What Do the Markings on Circuit Breakers Mean?

by Ed Larsen, P. E.

Have you ever been confused about what the markings on circuit breakers mean? Understanding the markings on electrical equipment is a fundamental need to ensure a safe and reliable electrical installation. Circuit breaker marking requirements are established by the requirements found in the *NEC* and the UL 489 product standard. This article will discuss the most common markings and where they can be found.

The UL 489 product standard for Molded Case Circuit Breakers specifies the information to be marked on circuit breakers and where it is to be located, so let's discuss what information needs to be marked on the circuit breaker and the location where you will find those markings. Keep in mind the UL® standard specifies minimum requirements. Circuit breaker manufacturers may provide additional information or provide information in a more convenient location.

Markings Visible without Removing Trims or Covers

UL 489 requires that some markings be visible without removing trims or covers. This location is typically referred to as the handle escutcheon (see photo 1).

Markings Visible with Trims or Covers Removed

UL 489 requires other markings be visible on an installed circuit breaker with trims or covers removed. This location is typically referred to as the face of the circuit breaker (see photos 2–4).

Other markings which should be visible with trims or covers removed are:

Independent trip – Multi-pole circuit breakers are constructed with either a common trip, where all poles are mechanically tripped when one of the poles trips, or an independent trip construction where only the pole that is involved with the overcurrent condition trips. If a 2-pole circuit breaker does not have an internal com-

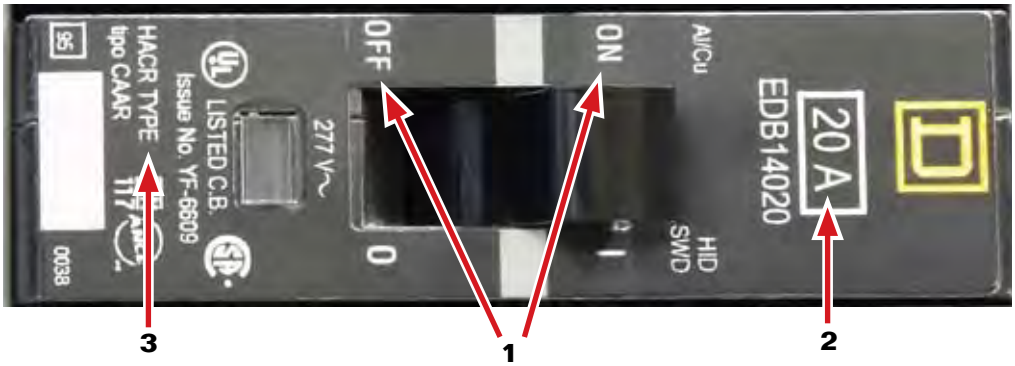

Photo 1. Markings Visible with Trims or Covers in Place

1. ON and OFF – The ON and OFF (closed and open) positions of the handle must be marked (*NEC* 240.81). These positions may also be marked with the internationally recognized "I" and "O" symbols, although this is not a UL requirement. If these markings are not visible when a motor operator is installed over the circuit breaker markings, then they must appear on the motor operator. Motor operators may be found in applications where remote or automatic operation of a circuit breaker is required.

2. Ampere rating (if 100 A or less) – The ampere rating may be located on the handle escutcheon or on the handle itself (*NEC* 240.83(B)). Circuit breakers that are rated more than 100 A may have their ampere rating marked in a position that is not visible with trims or covers in place.

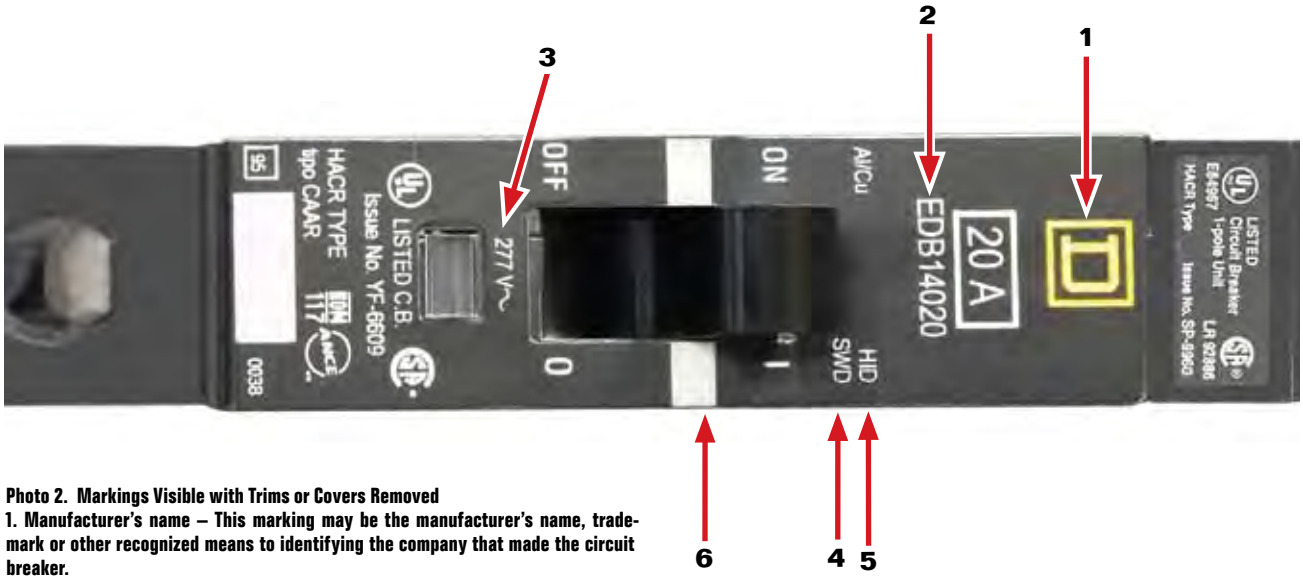

Photo 2. Markings Visible with Trims or Covers Removed

1. Manufacturer's name – This marking may be the manufacturer's name, trademark or other recognized means to identifying the company that made the circuit breaker.

2. Type designation – All circuit breakers are marked with a type designation, which may be a catalog number prefix or a separate designation. Equipment labels, such as on panelboards, will list the circuit breaker types suitable for use. Note that the word "type" may or may not be used on the circuit breaker or equipment labels. It is important to review the markings on the equipment, such as a panelboard, to make sure the circuit breaker designations on the equipment match the marking on the circuit breaker.

3. Voltage rating – All circuit breakers must be marked with a voltage rating. If the rating is not marked "ac" or "dc," then it is suitable for both. 120/240 V rated circuit breakers are suitable for use on single and three-phase 4-wire systems where the line-to-ground voltage does not exceed 120 V. Wye rated circuit breakers such as those rated 480Y/277 V, are suitable for use on three-phase 4-wire systems where the voltage to ground does not exceed 277 V. Special attention needs to be given to high leg or corner-grounded delta systems to insure that the circuit breaker has the appropriate rating. A review to see that the circuit breakers installed have a voltage rating suitable for the application is paramount for a code-compliant installation (*NEC* 240.83(E)).

4. SWD – 15- or 20-A circuit breakers rated 347 V or less may be marked "SWD," meaning that they are suitable for switching fluorescent lighting loads on a regular basis (*NEC* 240.83(D)). These circuit breakers are evaluated for high endurance use, since they will be used similar to a light switch.

5. HID – 50 A or less circuit breakers rated 480 V or less may be marked "HID," meaning they are suitable for switching high intensity discharge or fluorescent lighting loads on a regular basis. These circuit breakers may employ a different construction than a standard SWD circuit breaker in order to address the high inrush current resulting from the lower power factor created by the HID lighting (*NEC* 240.83(D)). These circuit breakers also undergo additional endurance evaluation to demonstrate their ability to perform the switching duty.

6. Trip and reset – Circuit breaker handles typically assume an intermediate position when tripped. This position must either be marked on the circuit breaker or on the equipment into which it is to be installed. If these markings are not visible when a motor operator is installed, then a "tripped" marking may appear on the motor operator.

mon trip feature, then it must be marked "Independent Trip" or "No Common Trip." *NEC* 240.20(B) is the foundational requirement for a common trip function in a circuit breaker; however, it also goes on to explain where independent trip is permitted.

For Replacement Use Only not-CTL – The Class CTL (circuit limiting) panelboard has only been in existence for about 25 years, even though the lighting and appliance branch circuit panelboard has been in the *NEC* for decades. CTL panelboards have a rejection means designed to reject more than the appropriate number of circuit breakers that can be installed in the panel. The marking "For replacement Use Only Not CTL Assemblies" means that the circuit breaker does not have CTL rejection provisions and is intended for replacement in older equipment pre-dating the CTL requirements for circuit breakers and panelboards. Circuit breakers with this designation should not be installed in a panelboard marked "Class CTL Panelboard" since that would be a violation of the listing of the assembly [*NEC* 110.3(B)].

Markings Found in Other Locations

The markings we will discuss below may appear in any location except the back of the circuit breaker. These markings include:

40°C – This marking indicates the maximum ambient temperature in which the circuit breaker can be applied at its marked ampere rating without rerating the ampacity of the circuit breaker. This marking is required for thermal-magnetic circuit breakers and is optional for electronic trip circuit breakers unless they are only suitable for a 25°C ambient, in which case they must be marked 25°C. When the ambient temperature rises above 40°C, the designer may need to consult the manufacturer to obtain rerating information (see item 4 in photo 3).

Class CTL – Circuit breakers marked Class CTL have a rejection means designed into the circuit breaker. Class CTL panelboards or assemblies, in conjunction with Class CTL circuit breakers, prevent more circuit breaker poles from being installed than the number for which the equipment is rated.

Photo 3. Markings Visible with Trims or Covers Removed

1. Line and load designation – Circuit breakers marked with "line" and "load" designations are not suitable for reverse connection. Circuit breakers with interchangeable trip units must be marked "line" and "load" unless there is no risk of shock when changing the trip unit.

2. Interrupting ratings – All circuit breakers with an interrupting rating more than 5000 A must be marked with an interrupting rating (*NEC* 240.83(C)). Interrupting ratings are stated in RMS symmetrical amperes. If the short-circuit current rating of the equipment in which the circuit breaker is installed is less than the interrupting rating of the circuit breaker, then the lesser rating applies. Circuit breakers should be reviewed after installation to ensure they have an interrupting rating suitable for the application. This marking may be found in any location except the back of circuit breakers that are 1-½ inches wide per pole or less due to the size constraint.

3. Ampere rating (if more than 100 A) – The ampere rating of a circuit breaker larger than 100 A may be found in a location that is visible after the cover or trim is removed. This marking requirement also applies to interchangeable trip units (*NEC* 240.83(A)).

5. Terminations (Cu-Al wire) – Circuit breakers must be marked with the type material (Cu-Al) and size of wire for which their terminals are suitable for use. If suitable for use with only copper or only aluminum, then the word "only" must be used. The abbreviations "CU" and "AL" are generally found on circuit breakers as permitted by the product standard. If only solid 10–14 AWG wire can be used, then that information must be noted. The number of wires per terminal will also be noted if more than one wire per terminal is permitted (*NEC* 110.14(A)).

6. Tightening torque – The nominal torque for all field-wiring terminals must be marked. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be found in any location except on the back.

7. Wire temperature ratings – Circuit breakers rated 125 A or less may be marked as suitable for use with 60°C, 60/75°C or 75°C only wire. Circuit breakers rated more than 125 A are rated for use with 75°C wire; the marking is optional. It is always permissible to use wire with a higher temperature rating, but it must be sized in accordance with the temperature marking on the circuit breaker and *NEC* Table 310.16. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be in any location except on the back (*NEC* 110.14(C)1).

8. HACR type – This marking indicates the circuit breaker is suitable for use with the group motor installations typically found in heating, air conditioning and refrigeration equipment. The *NEC* 2005 no longer has this marking requirement. The electrical industry determined that circuit breakers are considered suitable for use with such equipment without any further testing, therefore, the HACR marking is no longer required on air conditioning and refrigeration equipment or on circuit breakers for use in these applications. The requirement for this marking has also been removed from the UL 1995 product standard for HVAC equipment (see item 3 in photo 1).

Maximum wire size – Circuit breakers are typically marked with a wire range, however that marking is not mandatory. If the circuit breaker cannot accept the next larger wire size required for the ampere rating, then the maximum wire size must be marked in any location except the back (see item 5 in photo 3).

Separately shipped connectors –If connectors are not factory installed on a circuit breaker, then it must be marked with the proper connectors or terminal kits required in any location except the back (see item 8 in photo 3).

Ground-Fault Protection for People

The GFCI function, as part of a circuit breaker, provides ground-fault protection for people and has a number of unique marking and instruction requirements.

Test function – The GFCI has a test function that requires action upon installation and on a monthly basis. GFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate monthly testing.

Photo 4. 1Ø – 3Ø Marking
1. 1Ø – 3Ø marking – A 2-pole circuit breaker used to protect a 3-phase load on a corner-grounded delta system must be rated and marked for such an installation. Circuit breakers marked "1-phase – 3-phase" or "1Ø – 3Ø "are suitable for use on 3-phase corner-grounded delta or single-phase circuits (*NEC* 240.85).

Suitable for continuous operation at 100% of ampere rating only if used in a recognized enclosure 52 in. (1321 mm) by 20 in. (508 mm) x 7 1/2 in. (190 mm) deep or larger. Use only with 90° rated wire sized per the ampacity of 75° rated wire.

Figure 1. 100% Rated Marking
100 percent continuous rated – Circuit breakers are typically intended for use at not more than 80% of rated current where the load is considered continuous, or will continue for 3 hours or more (*NEC* 210.20). However, some circuit breakers are rated for continuous use at 100% of their current rating. These circuit breakers must be so marked in any location except on the back. Enclosure information such as a specific type or specific volume must also be marked. A requirement for the use of 90°C insulated wire sized to the 75°C column in *NEC* Table 310.16 and specific ventilation requirements may also be marked on the circuit breaker or equipment (*NEC* 210.20(A) and 215.3).

"Class A" marking – A "Class A" ground-fault device is intended to protect people. The Class A marking indicates that the trip threshold of the GFCI is between 4 mA and 6 mA. This marking may be in any location except the back.

Instructions – All GFCI circuit breakers must include instructions for the installer plus instructions on the use of the test function. A hangtag or self-adhesive label must also be provided, instructing the user to test the GFCI at least monthly. Inspectors should check to see that the tag or label has been properly installed.

Ground-Fault Protection for Equipment

Circuit breakers may also include a ground-fault protection for equipment (GFPE) function that, like GFCIs, has a number of unique marking and instruction requirements.

Test function – GFPE circuit breakers may have a test button or switch that may be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Trip level – GFPE circuit breakers must be marked with their trip threshold in milliamperes in a location accessible without removing trims or covers.

Instructions – All GFPE circuit breakers must include instructions for the installer.

Arc-Fault Protection

Circuit breakers may also include arc-fault protection (AFCI) that, like GFCIs, also has a number of unique marking and instruction requirements.

Device identification – AFCIs must also be identified appropriately. Branch/feeder or Combination type AFCIs must be so marked in a location visible when the trims or covers are removed. This is an important marking to note as we move into 2008, as *NEC*-2005 requires Combination AFCIs in bedrooms effective January 1, 2008 (*NEC* 210.12).

Test function – AFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Instructions – All AFCI circuit breakers must include instructions for the installer.

Circuit Breaker Markings Ensure a Safe Electrical Installation

So why are all of these markings on circuit breakers? Without them, it would be nearly impossible to install or inspect an installation for the appropriate performance ratings and fundamental electrical connections. When designing or completing an installation, key items to review are:

1. Are the voltage, continuous current, and interrupting ratings appropriate for the application?
2. Does the application require SWD or HID ratings?
3. Is the wire type and size appropriate for the circuit breaker?
4. Is the circuit breaker suitable for the equipment in which it is installed? Have other protective functions such as GFCI or AFCI been provided as required by the *NEC*?
5. Is the temperature rating of the circuit breaker suitable for the application?

The UL Marking Guide for Molded Case Circuit Breakers is a valuable resource to understand circuit breaker markings that may further explain these and other markings in detail. If you have questions about CB markings not answered here, consult the Marking Guide or the manufacturer to assist in an *NEC*-compliant installation.

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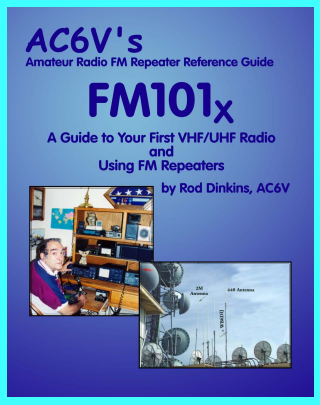

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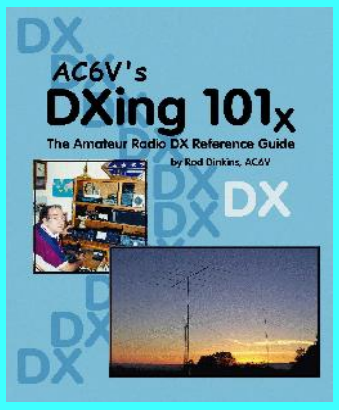

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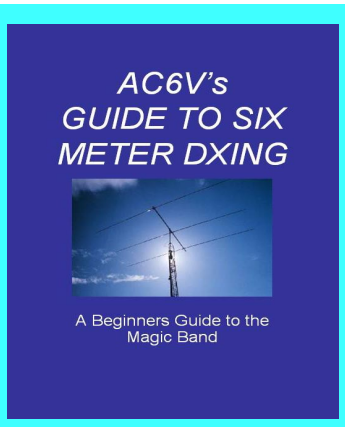

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J-Poles -- From The ARRL
Super J-Pole Antenna(Collinear Design) - By KB0YKI 2M, 220, 440, 6M
J-Poles Again -- From Stephen R. Yates, AA5TB
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Log Periodic Antenna Calculator
Long Wire Antenna
Long Wire Impedance matching
Full Wave Loop Antenna 10M thru 80M -- From Western Canada's Ham Radio YL Site
Full Wave Loop Antenna -- From WH2T Via KQ8RP
Loops -- From The ARRL
Loop Antenna for HF Receiving -- From K7ZB
Loop Antenna - Low Noise Coax for 160 - 10 meters -- From WN6F
Loop Antennas -- Plans for each Amateur Band
Loop Antenna Forum -- Yahoo -- Needs subscribing but lots of info here

Loop Antenna, Magnetic, Portable -- From G4FON
Low Band Antennas -- From K3KY

Mac Antenna Master Software
Magnetic Loops for HF -- From W2BRI
Magnetic Loops -- By Wolfgang DJ3TZ
Magnetic Loop Antenna -- From HB9ABX
Magnetic Loop Design -- From QRZ
Magnetic Loop Antennas By GM3MXN
Magnetic Loop Antenna's
Microwave Antenna HandBook -- From W1GHZ
Mobile Antennas -- From The ARRL
Mobile Dual Band VHF/UHF Antenna -- Homebrew Project from VE3RGW
Mobile Antenna Motorized, 98% plastic 5 ft long -- From VE3GK
Mobile HF Antenna By HB9ABX
Moxon Antennas Via L. B. Cebik
Moxon Antenna Projects -- 17M and 20M Via KD6WD

About Mobile Dual-Band 2m/440 -- Roger K6XQ advises - Can I build a 2m/440mhz 5/8 mobile antenna? A 1/4 wave on 144 MHz is 3/4 wave on 440. The SWR is usually acceptable on the 3rd harmonic, although that is partly due to the increased feeder losses on 440. But SWR is not the whole story. A 3/4 wave radiator on 440 has radiation lobes that send some of your signal at high angles, not toward the horizon where you probably want it. This is usually not so severe to make the antenna unusable, and may be considered a reasonable compromise for the dual band coverage. A 5/8 wave antenna on 144 MHz is 15/8 waves on 440. If you do the simple math you will see that on 440 you will have almost a 2 wavelength antenna, which is becoming what we call a "long wire" on the HF bands. As you probably know, a long wire radiates with multiple lobes, the largest of which are toward the ends of the wire. In other words, most of your signal will be radiated up toward the sky. The reason for the coil in commercial dual band antennas is to phase the antenna currents so you actually end up with two colinear elements which both radiate at low angles, reinforcing each other and providing gain. Without the phasing coil you get multiple undesirable lobes. Also keep in mind that 5/8 wave is NOT a resonant length. There are several ways to compensate for the reactance at the base of a 5/8 wave radiator, the easiest being to add a small inductance in series which makes the antenna electrically equivalent to a 3/4 wave. (3/4 wave is resonant, as you have already discovered). Given that the gain advantage of a 5/8 wave antenna over a 1/4 wave antenna is probably not noticeable except under the most difficult conditions, and even then may or may not make a difference, my suggestion is to stick with the 1/4 wave antenna. It is simple and works OK.

NVIS -- Near Vertical Incident Skywave Antennas

Petlowany Three-Band Burner Antenna -- Trapless short vertical antenna - Resonant on 20, 15 and 10 meters, without traps, 12 and 17M with a tuner. From David, N5IZU
Pfeiffer Quad -- From kQ6RH
Phasing Arrays - Vertical Antennas -- From Butternut
Portable Antennas -- From The Ham Club At University of Hawaii at Manoa
Portable Antennas -- From HFPack
Portable dual-band antenna for 20 and 10 meters. - The Flower Pot Antenna from IZØFYL

QRZ Shareware Collection antennas -- over 40 antenna files
Quad Antenna Calculator -- From KD6DKS
Quad Antenna Design -- By: Dr. Carl O. Jelinek N6VNG

Quads By EI7BA
Quads -- From KQ6RH
Quad Antennas --- From The ARRL
Quads 10-12-15 Meters From IZ7ATH
Quad Loops by Ray Jurgens KQ6RH
Quad - Multiband - By EI7BA
Design-A-Quad Antenna Via K4ABT
Quad loops - From KQ6RH

Reflected M beams by Ray Jurgens KQ6RH
Rhombic Antennas 30M - 6M -- From KC0FVV
Rhombic Antennas -- From Integrated publications
Rhombic Antennas -- From Ian Cummings

Roof Top Tower -- By KB0YKI
Rotary Dipole For 17 And 20
Rubber Duck Antennas

Satellite Antennas
Shortwave Receiving Antennas
Skywire Loop Antenna -- From The ARRL
Slinky Antennas -- From Antennas and More
Sloper Antennas -- From The ARRL
St Louis Vertical -- From The The American QRP Club
Stealth Antennas
Sterba Curtain Antennas
Sterba Curtain Antennas -- From Ham Universe
Sterba Curtain 40M -- from KB8PGW

T2FD Antenna
T-Hunt Antennas
Telrex Antenna Trap Repair -- From N6KI
Transmission Line Calculator -- From AA3RL
Traps From VE3GK
Trapped Antennas -- From The ARRL
Turnstile Antennas -- From Integrated publications

Tower Talk Archives -- Lots of advice from a reflector
Tower Installation -- Photos From W5AJ
Tower Installation -- Loaded With Info -- From N1LO
Tower Installation -- See how KO4BB did it

Tower/Antenna Raising -- From Dr. Dave
US Tower Specs, Base Data & More
Titan Tower Specs, Base Data & More Via Champion Radio
The Ten Most Common Tower Building Mistakes
MARC - The Mast, Antenna, And Rotor Calculator Via Champion Radio

Vee Antennas -- From Integrated publications
Inverted Vee And Dipole Calculator
Inverted Vee - Linear Loaded By KGØZP
Vertical Antennas -- From The ARRL
Vertical Wide Band 7-21 MHz -- The RXO Unitenna -- from G3RXO
Vertical Antennas - Dirty Little Secrets, Ground/Radial Systems, Phased Arrays of Short
Vertical Antennas, Why Radials?- A MUST READ - From Butternut antennas Click On Tech Notes
Vertical Antenna - transportable By CT1BYR
Vertical Antenna Loading Coils -- see LOADCOIL * Design short vertical antenna and loading coil. From Reg, G4FGQ
Vertical Antenna Phasing Arrays - From Butternut Antennas
Short Vertical Antennas and Ground Systems - VK1BRH
Vertical - Quickie Style -- From KQ6RH
Vertical Antennas & Ground Screens -- By N6RK
Vertical Dipole -- from Tom Sevart, N2UHC
Vertical Dipole -- From L. B. Cebik, W4RNL
VK2ABQ Wire Beams by Ray Jurgens KQ6RH
VHF Antennas - Various -- From The ARRL

W5QJR Antenna Via Antennex
Water Antennas -- Not April 1 stuff
WaveGuide Tin Can Antennas -- From Gregory Rehm
Windom Antenna -- From JA7KPI
Windom Antenna -- From BuxCom
Windom - 6 Bands - From K3MT and daughter KF4LGR
Wire Antennas -- Raising and building -- From N1LO

X-Beams -- By KQ6RH

Yagi -- OWA -- Optimized Wideband Antenna -- Follow On Page By NW3Z
Yagi Optimized Wideband Antennas By NW3Z
Yagi Modeling -- Six Antenna Comparison By L. B. Cebik, W4RNL
Yagi/Beams -- From The ARRL
Yagi - Optimized Wideband Antenna (OWA) and Skyhawk, a state-of-the-art trapless tribander

Yagi Designs - Lots and Lots -- From TEARA - Click on Knowledge Quest, then topic.

Stacking Yagis -- From The ARRL

Wire Yagis -- from VE7CA

Double Extended Zepp Dipole -- Cut for 15 Meters
Zip Cord Wire As Transmission Lines and Radiators -- From The ARRL -- Click On Other HF Antennas
ZL Special -- From The ARRL -- Click On Other HF Antennas

160 Meter Antennas

160M/80M Coaxial Receiving Loops -- By KC2TX
160M Super Linear-Loaded Inverted V By KGØZP
Balloon and Kite Antennas On The Top Band -- From G4VGO
Battle Creek Special -- Info From PI4CC Contest Club
K5OE Antennas -- HF Antennas 40/80/160 55 ft Vertical
KGØZP Super Linear-Loaded Inverted V
Low Band Antennas -- From K3KY
Verkürzte Antennen -- Short Dipoles and Verticals for 160m & 80m -- From DJ9RB
Skywire Loop Antenna -- From The ARRL

80 Meter Antennas

Antenna Projects From The YCCC -- Double-L For 80/160, Two Wire Beverage, A Poor Man's 160 Meter 4-Square, Using a 4 square Vertical Phased Array to improve your 80 and 160 meter signal, -- From The Yankee Clipper Contest Club
Verkürzte Antennen -- Short Dipoles and Verticals for 160m & 80m -- From DJ9RB
160/80/40 Meters - 55-Foot Vertical From K5OE
VE3GK's Homebrew Site -- 80M Rotating 2 element Quad
Verkürzte Antennen -- Short Dipoles and Verticals for 160m & 80m -- From DJ9RB
40M/80M Trap Dipole -- From QRZ
80/40 Meter Vertical From EI7BA
80 Meter Vertical (Half length) Monopole From L. B. Cebik, W4RNL
80 Meter Inverted Vee From GERRY VE3GK
80/40 Super Loop Antenna
80 Meter Antenna -- Reduced Size For Small Lots

2 EL SHORT BOOM 80M YAGI From VE6WZ.
Short Dipole for 80M -- From 4S7NR
80 Meter Frame Antenna -- From Harry Lythall - SM0VPO
Low Noise Coax-Shielded Faraday Loop Antenna For 40M and 80M
Full Wave Loop Antenna 10M thru 80M -- From Western Canada's Ham Radio YL Site
Taylor Space Miser 80M Antenna

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40 Meter Antennas

40M Shortened Loop -- By Ben Smith, W4KSY Via Antennex
WB0NNI 40 Meter Linear Loaded Vertical
K5OE Antennas -- HF Antennas 40/80/160 55 ft Vertical
St. Louis Vertical -- By Dave Gauding, NF0R
DDRR - for 40M -- Directional Discontinuity Ring Radiator Antenna
40M Beam -- From QRZ
40M Delta Loop
40M/80M Trap Dipole -- From QRZ
40 Meter Delta Loop By PY4VE Via K4TX
Stealth 40M Antenna - From K7ZB
ShortyForty Dipole -- From FlashWebHost
Sterba Curtain 40M -- from KB8PGW
Low Noise Coax-Shielded Faraday Loop Antenna For 40M and 80M

20M Antennas

20M 3 Element Monobander -- From The ARRL
K5OE Antennas -- HF Antennas - 10/15/20 m Dipoles
VE3GK's Homebrew Site -- 7Element 20M Yagi 63ft Boom
20 meter Extended Double Zep From NB6Z
20M Yagi Optimized Wideband Antenna
20M Delta Loop
Moxon Antenna Projects -- 17M and 20M Via KD6WD
Verticals

17M Antennas

17M Beam Antennas -- From The ARRL
Moxon Antenna Projects -- 17M and 20M Via KD6WD
Verticals

15M Antennas

15 Meter Beam -- From The ARRL
K5OE Antennas -- HF Antennas - 10/15/20 m Dipoles
15M Delta Loop
15M Yagi Optimized Wideband Antenna
A One Element V Beam For 15 M From KB4XJ
Verticals

12M Antennas

12M Beam Antennas -- From The ARRL
Verticals

10 Meter Antennas

Two element 10-Meter beam -- From The ARRL
10 Meter Verticals using modeling -- From NM5K
10M 3 Elements By DF9CY
10M Yagis a Bunch -- From The Antenna Elmer
10M Yagi Optimized Wideband Antenna
10M Vertical Dipole -- from Tom Sevart, N2UHC
10M Vertical Dipole -- From L. B. Cebik, W4RNL

HF TRI-BANDERS & MULTIBANDERS

5 Band Log Periodic Dipole Array -- From VE7CA
12- and 17-meter lightweight Yagi -- From The ARRL
30-17-12 Meter Moxon Rectangles - From L. B. Cebik
All-Band Center-Fed Inverted-L -- From L. B. Cebik
All-band Wire Antenna - From Joe Tyburczy, WB1GFH
MULTIBAND HF DIPOLE PROJECT
ONE ELEMENT BEAM 20 THRU 6

Cobwebb HF Antenna -- 14 - 28 mhz
Double-L Antenna For 80/160 -- Via The YCCC
Full Wave Loop Antenna 10M thru 80M
G5RV Multi-Band Antenna by Louis Varney

IK-STIC -- multi-band vertical dipole antenna used for quick set-up and quick band change -- From W2IK
IK-STIC 2 vertical, all band, antenna, 25 feet tall, under 5 pounds! With a tuner covers the amateur radio HF bands from 40 - 10 Meters.

Lattin 5 Band Antenna -- BY W4JRW Via G3YCC
Multiband Antennas -- From The ARRL
Multiband Dipole Antenna -- From The N4UJW Antenna Design Lab
MultiBand Quad - By EI7BA
MultiBand Vee Beam -- Six Band One Element Beam 20 meters thru 6 From LA0HV
Pedestrian 5-Band Mobile Antenna - From W3FF
Petlowany Three-Band Burner Antenna -- Trapless short vertical antenna - Resonant on 20, 15 and 10 meters, without traps, 12 and 17M with a tuner. From David, N5IZU
Portable All Band Antenna -- "IK-STIC" -- a multi-band vertical dipole antenna which can be used in the field for quick set-up and quick band change.
Quads 10-12-15 Meters From IZ7ATH

Rotary Dipole FOR 17 AND 20
Spider Beam Portable Triband Yagi -- 10/15/20M Via Con DF4SA
Sturba Curtain - All Band
St. Louis Vertical -- From Dave Gauding, NF0R
Tri-Band 2 Element Portable Yagi -- From Markus Hansen, VE7CA
Two Element Beam I for 10-, 15-, or 20-meters -- From The ARRL
Two Element Beam II for 10-, 15-, or 20-meters -- From The ARRL
Vertical Wide Band 7-21 MHz -- The RXO Unitenna -- from G3RXO
Windom - 6 Bands - From K3MT and daughter KF4LGR
Windom Multibander -- From BUXCOM

6 Meter Antennas

5/8 Six Meter Vertical -- by G3JVL
50 Mhz Antennas By ON4ANT
Six Meter Antennas A Bunch -- From The UKSMG News
Six Meter Antennas A Bunch -- From 6MT.com

Copper Cactus J-Pole antenna -- Plans for 52MHz -- By KG0ZP
Six Meter J-Pole -- By K4ABT
Super J-Pole Antenna(Collinear Design) - By KB0YKI
Simple Six Meter Dipole Horizontal or Vertical -- By JIM BAUDO

Six Meter Halo -- From Steve KB1DIG
Six Meter Halo -- From W3DHJ
Six Meter Sloop By K0FF
Six Metre Long-wire Aerial -- By Brian D. Williams, GW0GHF

Six Metre Indoor Loop -- By Colen Harlow, G8BTK
Six Metre Antenna -- By Maurius - ZR6YY
Six Meter Base Station Antenna -- By K4ABT
Six Meter Long Wire Antenna - From GW0GHF
Moxon Rectangles for 6 Meters -- From W4RNL, L.B. Cebik

6M and 2M Yagi -- From The Antenna Elmer
6M Yagis a Bunch -- From The Antenna Elmer
Six Meter Yagi -- From The ARRL
Six Meter three-element Yagi -- From The ARRL
Six Meter Tri-Yagi -- From The ARRL
50 MHz 6 Element Beam By DF9CY
Six Meter Monster Beam -- 8 element, 41 foot boom, 14.0 dBi gain, By N6CA
Optimised Six-Metre Yagi by Brian Beezley, K6STI
Two portable 6 meters antennas - two-element quad and a three-element yagi with telescoping elements -- From VE7CA

Verticals

2 Meter Antennas

2M/440 MHz Dual Band Copper J-Pole From N7QVC

VHF/UHF Antennas -- By KB0YKI -- J-Poles, Super J-Pole, Omnidirectional, Quads, Yagis
2 Meter Antennas A Bunch -- From 6M.com
American Legion J-Pole - Portable roll-up Antenna
BiQuad For 2m/440 - FROM KE4UYP
6dB COLINEAR VHF ANTENNA -- From Harry Lythall - SM0VPO
The Simplest Collinear -- From Ross W1HBQ
DDRR - for 2M -- Directional Discontinuity Ring Radiator Antenna
Cycloid Collinear, a CP Omni for 2M from Ross W1HBQ
HO Collinear, a Horizontal Omni for 2M from Ross W1HBQ

Copper Cactus J-Pole antenna -- Plans for 52, 146, 223.5, 435, 912, and 1265MHZ -- By KG0ZP
2M Fan Antenna (Yes From an Electrical fan)
Two Meter Hanging Dipole - From Peter Parker VK3YE
Grid Yagi -- From Ross W1HBQ
2M Half Square Antenna With PVC Support -- Via L.B. Cebik W4RNL
Halo Antenna 2M-- From N2KBK
Halo Antenna 2M Square -- From N4UJW
Stacked J-Pole Plans I
Stacked J-Pole Plans II
2M Twin lead J-Pole By K4ABT
2 M 1/2-Wave J-Pole -- VHF-FM (Stealth) Antenna Made From 450-Ohm Ladder Line -- From KB1GTR & KB1DIG
2M Log-Periodic Dipole Array -- From The ARRL
2 METER SSB SQUARE LOOP From N4UJW
N2NJH's Antenna Roundup -- Build a J-Pole for any frequency - Includes an Excel Spreadsheet!, Pocket J-Pole, Half Wave J-Pole, Stacked 5/8, Field 2 meter Colinear, 2 meter Colinear in PVC, 2 meter Halo, Eggbeater Loop Antenna
PacketRadio Operator's Antenna Handbook -- Antenna Basics, 2M & 6M Antennas
2m Quagi By W5UN
2M Quadix -- from Ross W1HBQ
SlimJim 2M Antenna
Sperrtof 2M Antenna -- aka Sleeve Dipole or Sleeve J-Pole. By ON4CFC Via Antennex
Super J-Pole Antenna(Collinear Design) - By KB0YKI
VHF Magnetic loop By ON1DHT
Cheap Yagi Antennas for VHF/UHF -- From The Clear Lake Amateur Radio Club
3 element 2-meter Yagi -- From The ARRL
5 element 2-meter Yagi -- From The ARRL
2M 7 Element Yagi
2M 9 Element Yagi
2M Yagis a Bunch -- From The Antenna Elmer
2M Yagis. Old TV Antenna Scheme. Submitted by Gary Ruehle. Drive around and find/collect old TV antennas that people are throwing away. My son and I built a dual 6 element stacked yagi for 2 mtr a couple of years ago. TV antennas were all the building materials we used. After the antenna was complete, we could hit 2 different repeaters about 60 mi away using a 1 watt handheld. This antenna is highly directional.

220 MHZ Antennas

Cheap Yagi Antennas for VHF/UHF -- From The Clear Lake Amateur Radio Club
220 MHz Quickie Quarter Wave Antenna -- From Artsci
220 MHz J Pole Antenna -- From N6ZAV
Super J-Pole Antenna(Collinear Design) - By KB0YKI

440 MHZ Antennas

Cheap Yagi Antennas for VHF/UHF -- From The Clear Lake Amateur Radio Club
Copper Cactus J-Pole antenna -- Plans for 435MHz-- By KG0ZP
Coaxial Collinear Antenna -- 432 MHz -- From Via N9ZIA
Small 70cm Yagi -- From The ARRL
9 dB, 70cm, Collinear Antenna From Coax -- Version I -- From N1HFX

2M/440 MHz Dual Band Copper J-Pole From N7QVC

432 8 Element Quagi Antenna By N2KBK
440 MHz -- 2 Element & 4 Element Yagi's - From The Antenna Elmer
430 MHz Antennas A Bunch-- From 6MT.com
Coat Hanger 7dB, 4-El 70cm Yagi -- Simple, cheap, RUGGED & unscrews for field work--via Stan ZL2AJZ
Super J-Pole Antenna(Collinear Design) - By KB0YKI

900 MHz Antennas

900 MHz Antennas A Bunch -- From 6MT.com
Cheap Yagi Antennas for VHF/UHF -- From The Clear Lake Amateur Radio Club
Copper Cactus J-Pole antenna -- Plans for 912MHz - By KG0ZP

GHz Antennas

GHz Antennas A Bunch -- From 6MT.com
Copper Cactus J-Pole antenna -- Plans for 1265MHZ -- By KG0ZP
Cheap Yagi Antennas for VHF/UHF -- From The Clear Lake Amateur Radio Club
1270Mhz Yagi Antenna
Microwave Antennas and Greg's Wireless Networking Info Page
2.4GHz Omni Antenna
2.4 GHz Sardine Can Antenna -- A Double Quad Bow Tie
Loop-Uda-Yagi para 2.4 Ghz.

STEALTH ANTENNAS

THE DREADED CC AND R's

CC&R INFO

Condo Communicator -- Help for Condo and Apartment Dwellers -- From WA5OES

CC&R's? (You Got) -- Info from the ARRL

K3QK's Legal Resources -- For hams can use it for that "tower project"

eHam.net CC&R Survival Series

eHam Forum on Antenna Restrictions

FCC Fact Sheet - Over-the-Air Reception Devices Rule Preemption of Restrictions on Placement of Direct Broadcast Satellite, Multi-channel Multi-point Distribution Service, and Television Broadcast Antennas.

Antenna Restrictions? -- From AC6TS

Antenna Restrictions - How To Chart - From The ARRL

Antenna Zoning for the Radio Amateur - ARRL Book On legal aspects of antenna restrictions

HOMEBREW STEALTH ANTENNAS

Balcony Antenna -- From Harry Lythall - SM0VPO

Hidden, Stealth, HF Antennas -- From K3MT -- The GRASSWIRE -- You read it correctly!!

Hidden, Stealth & Invisible Antennas -- Smartuners for Stealth Antennas From SGC, Inc.

Hidden Antenna Ideas -- Links Via AC6TS

Hidden Antenna Ideas Loops, Operation Away From Home, Attic, From N0HC

Stealth Amateur Radio - Book From The ARRL -- (ISBN: 0-87259-757-1) #7571

Stealth 40M Antenna - From K7ZB

RadialWave™ Ground Radial System -- Great for DEED RESTRICTED BASE STATIONS, QRP, BACKPACKING, FIELD DAY, OR EMERGENCY COMMUNICATIONS.

Stealth Antenna Experiences - From Clif's ham radio connection.

SGC Smartuners for Stealth Antennas -- Free - Click on "Download your copy today!"

10M Bent Attic Antenna - A Yagi No Less

Stealth 40M Antenna - From K7ZB

2 M 1/2-Wave J-Pole -- VHF-FM (Stealth) Antenna Made From 450-Ohm Ladder Line -- From KB1GTR & KB1DIG

Small Guage Wire Makes an "Invisible Antenna" The Wireman sells #26 AWG "Invisible Toughcoat Silky" antenna wire which is a stranded & coated copperweld type, his catalog #534. Amazingly strong for it's size, very abrasion resistant and you have to go looking for it if you're more than twenty feet away from the stuff.

COMMERCIAL ANTENNAS USED FOR STEALTH

BuddiPole -- W3FF Products
B&W Balcony/Window Antenna - AP-10A
CliffDweller II antenna
End-Fedz -- full length half wave dipoles, but with an important difference. The coax connector is at one end of the dipole, where it is most needed. From Universal Radio
EZ HANG -- SlingShot & Reel Device for installing wire antennas
Force 12- 9 Foot Vertical
Force 12 Aluminum Flag Poles & Antennas
HamSticks
Isotron Antennas
MFJ Apartment Antenna 40 Meters to 2 Meters MFJ-1622
MFJ Super Loop Antennas
MFJ ScrewDriver Antennas
Outbacker Antennas -- Terlin
Quicksilver Radio Products MinuteMan™ HF Portable Antennas
Stealth Antennas -- HF Mobile Antennas By VE7BOC
Texas BugCatcher - GLA Antennas
Texas Twister - GLA Antennas
Ventennas -- The Forbes Group
Ventanna HFP-2 10 Foot HF Antenna -- 6.5 to 30 MHz !!!
Vern Wright's HF Mobile Antennas, MP1
WeatherVane Antennas

SOME OF MY FAVORITE ANTENNA BOOKS -- AC6V

It Is Back In Print -- Ed Knoll's "Easy Up Antennas" - A good One

New From Ed Knoll "Dipole and Long Wire Antennas"

The ARRL Antenna Handbook Buy the Book Today

The Radio Amateur Antenna Handbook By William I. Orr Buy the Book Today!

Cubical Quad Antennas : How to Build and Adjust Quads -- by William I. Orr Buy the Book Today!

Beam Antenna Handbook by William I. Orr, Stuart D. Cowan (Contributor) Buy the Book Today!

Simple Low Cost Wire Antennae by Orr, William I. Orr Buy the Book Today!

Practical Antenna Handbook by Joseph J. Carr Buy the Book Today!

Antenna Toolkit by Joe Carr, Buy the Book Today!

Lew McCoy on Antennas Buy the Book Today!

ON4UN's Low Band DXing -- Great Source For Low band Antennas Buy the Book Today!

Go To The Software Pages -- For Antenna Simulation Programs

Commercial Antennas -- Visit AC6V's Antenna Mfgrs Links

AC6V's Amateur Radio Links **GO BACK TO THE BIG INDEX**

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_AC6V BRAG TAPE

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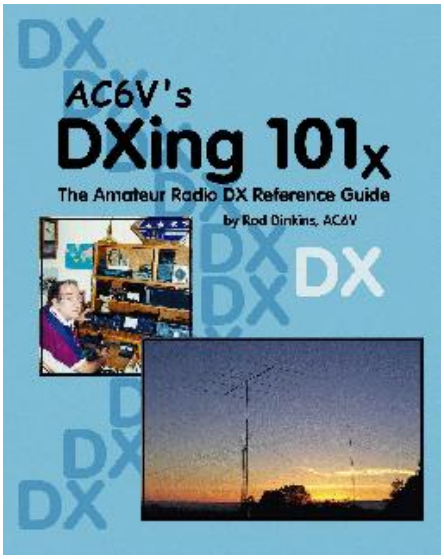
THE AMATEUR RADIO DX REFERENCE GUIDE

"Best Title On The Subject" -- Universal Radio



"This Book Has Been Needed For Years" -- World Radio Magazine

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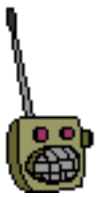

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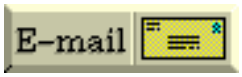

AC6V E-MAIL ADDRESS

This book is a year in the writing and features the DXing advice of several noted DXers and technical gurus. This is not a brag book on my personal DX exploits, but tried and true practical DX advice from those who have been there, done that, and worked them all.

It is intended for those entering into HFing and the wacky world of DXing. Old timers not yet on the Internet will want this book as well.

226 Pages

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DX101X is comb-bound so the latest updates will be inserted prior to all orders as we print on demand. This way all updated URL's, reader feedback, and corrections will be in your manual. As further addendums are received, they will be posted here under
MANUAL UPDATES

TESTIMONIALS (UNSOLICITED)

eHam Reviews -- 16 Reviews From DX101x Owners

World Radio Magazine DX101x Review -- Click Here

From Universal Radio Reynoldsburg, OH ***We are most impressed with the scope and currentness of this excellent book. Best title on the subject.*** **Universal Book Order Page**

GW0ANA wrote: Greetings Rod, I would like to thank you & yours first for your amazing book & secondly your taste in "Post Cards". I enjoyed looking at both.

Again I must applaud you for the immense effort you have put in on your book. The subject matter is of great help to Novice & Expert alike. It is like having an "DX Elmer" always at your shoulder.

I just love the relaxed way you impart real DX knowledge to the deserving. Your description on how you caught Malpelo was inspirational to this DX-ER. I got the guy on 10 by abiding by the first law of DX,"Listen & listen again" He popped up on 10 metres on a so called dead band, heard him & got him simplex 100 watts into a mini beam at 30 feet. I have big planning problems. But using wires in trees & lots of listening skill I have 310 countries confirmed & 5BDXCC. Again my thanks for your wonderful book I will dip in & out of it at will.

Best 73 Glyn, GW0ANA, Nirvana Castle Precinct, Wales UK
P/S I am sure your book is my "Passport to honour roll"

Hi Rod. I'm new to ham radio since late Jan of this year (KD7ZHS). I'm sure you've heard this 10,000 times, but your DX book has been an incredible help to me. Ditto for your web site. Hope to meet you on the air some day. Sincerely, Don Keller KD7ZHS

It's like, DXing for Dummies

Best book for DXing for newbies or perhaps anyone. I'm sure even the seasoned pro could glean stuff from here. It's a must have for people like me who know nothing. Well I know something now. This book is laid out logically and is an easy read. I even chuckled a few times. His sense of humor is perfectly timed. But mostly it's so handy and I keep referring too it daily while DXing. It's got useful tricks to DXing and tips on equipment and procedures. Really enjoyed his DXing secrets. IMHO, probably the most important thing is the frequency listings and their uses. I'm hoping that will help standardize things as far as where the DX is. But I'm amazed how knowledgeable AC6V is on this subject. I've got tabs pasted all over it for quick reference and it is right next to me in the ham shack. . . usually opened to something like frequency list or the "Q" codes. Tons of stuff in there like that. I'm VERY glad I bought this one and I don't buy many "how to" books. N5WVN

Fantastic dx book

What a great dx book! When I was called to jury duty I selected THIS BOOK to read while waiting in the jury assembly room! It's a great way to learn and review dxing techniques. Those dxing secrets are great (but don't tell anyone!). I know a lot of work went into this book and I thank and respect all those who worked on it. Another thing that makes it great is the humor mixed in - in fact, I'm asking my local radio club if they want to go on a dxpedition to Charcoal Island in the Barbecues IOTA group! Thank you Rod and everyone for a great book. 73 Martin G. (Marty) Blaise (AG5T)

Rod, 5 star rating. I just bought your DX 101 book, and although I already have snagged 225 countries and have DXCC, I have found quite a few really neat hints and tips that will prove useful. I must admit to reading a bunch of tips and thinking - "Hey that's right, that's what I do . . . ". So I'm learning new stuff and getting a kick out of the "secrets" that I already stumbled upon. Your book will be invaluable to beginners. I hope you can find a publisher who will bind using perfect bound or some regular book binding. . . Best DX and 73, Rich KY6R

Ed Note -- comb binding allows the book to be updated as additions and suggestions come in. These are incorporated into all new books -- books are printed on demand. Updates at URL:
MANUAL UPDATES

Although licensed for several years, been away from Amateur Radio for about 10 years, so getting restarted with all the new technology and information available was a chore, but DX101X made things a whole lot more enjoyable.

The bottom line is this, I have read the book three times and extracted only a small percentage of the available information; and, if you are either new to our hobby or think you have a strong enough ego not to be humbled but to absorb the information contained in this book, you are in for a very pleasant surprise! From KB1GZ at eHam.reviews

---

Hello Rod, Tks so much fer the book I rcvd hr last week – vy vy psed wid it es still a lot to learn/to pick up; in particular the contesting part. Cuagn 73 de Harry PA3ARM pa3arm@arrl.net

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Rod ... Outstanding book! Just got the book in the mail today. I've only had it a short time - but I can't put it down. This book is very well done - outstanding job! I can tell a ton of care and time went into this book. What I like the most about it is it is full of Practical information. I'm learning about stuff I've never really understood until I read this book. I hope you will keep adding to it an updating it. My two cents worth in just the first part of the book I've read - add the Icom - 746 to your list of good contesting HF rigs. I have the Yaesu 1000mp and I'd put the 746's receiver up against it any day. Keep up the great work and thanks for an outstanding book. Feel free to use my comments for your testimonials.73 Dave Sass KC0IWV, Savage MN kc0iwv@mindspring.com

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Hi Rod, Thanks very much for sending me an advanced copy of DXing 101x per Paul DeCicco's NN6X request. You've really done an outstanding job in writing a comprehensive text at a beginning and intermediate level for dx'ing. There's a great introductory section on contesting, one of my passions, that should serve the budding contester very well indeed. Included is a very complete treatment of all the important equipment features and performance parameters that the budding dx'er should give attention to when selecting his gear. You've put together in one place information that previously could only be gleaned from countless sources. It really is a unique and rich source of information! 73 Barry N1EU Delmar NY n1eu@yahoo.com

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Rod, All I can say about DX'ing 101X is it is Fantastic. An unbelievable reference for any dx'er. Thanks for your efforts. 73 de N1LQ, David Hammond, Middleton, MA dhhdeh@concentric.net

---

DX101 contains essential information for the beginner DXer. It is also valuable for the DXer who is seeking to tap the wealth of resources available on the Internet. I am neither and still enjoy reading thru this work. It is well written and deals with all relevant topics - it is a valuable DX tool ! Josh, Grand Junction CO, N7XM
n7xm@qsl.net

Rod, the book arrived and I haven't taken my nose out of it..... I know that a lot of hard work went into getting this book out ..... You would be surprised at all the helpful information that I've already acquired from reading your book.....and I have only gotten through chapter five..... Thanks to you and to all the people that assisted you in making it available to the Ham World.....Best regards..Bill NU5C, North Little Rock, Arkansas
Bdretpd@aol.com

Hello Rod, Just would like to congratulate you on the book. I'm just almost halfway reading it and I've already learned a lot. Considering that I just need a few countries to qualify for DXCC Honor Roll, I still find your book to be an invaluable tool for dx'ers. Congratulations! 73, de Ernie, DU1COO, also KH0DK
ariana@pacific.net.ph

MANY MORE UNSOLICITED TESTIMONIALS -- TOO MANY TO PUBLISH

SOME MORE REVIEWS AT eHam.net

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OTHER DX BOOKS

Low Band DXing By ON4UN URL: **http://www.universal-radio.com/catalog/books/3635.html**

The Complete DX'er By W9KNI URL: **http://www.idiompress.com/books-complete-dxer.html**

Actually you will need all three of these books to completely cover the subject of DXing: DX101, Low Band DXing, and The Complete DXer.

Thanks For Stopping by 73

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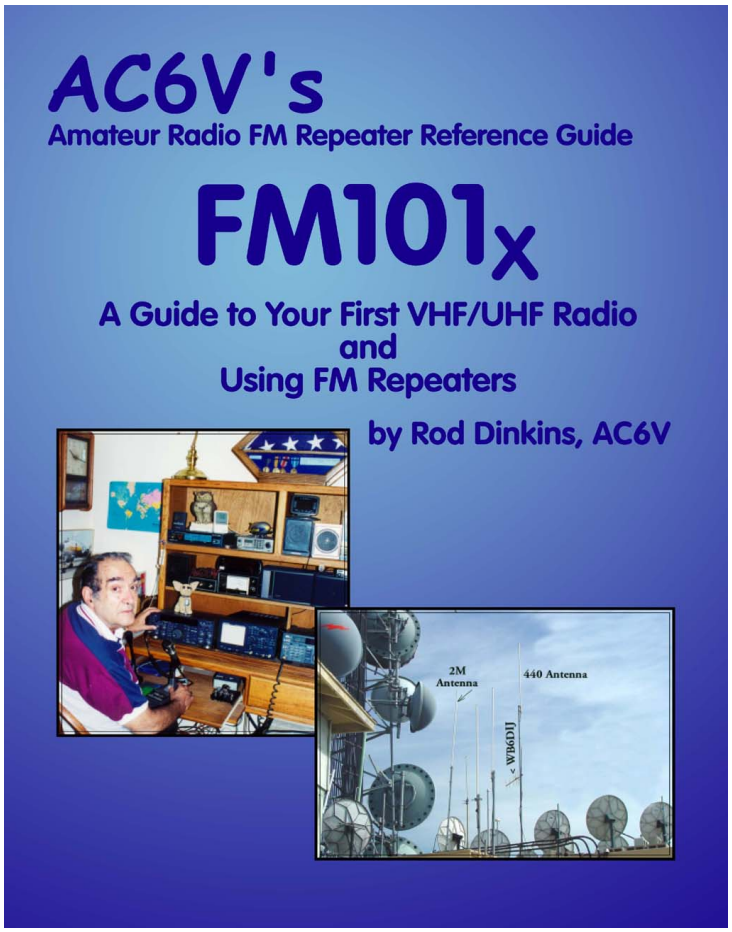
Updated Tuesday, August 31, 2004

AC6V's GUIDE TO FM REPEATERS



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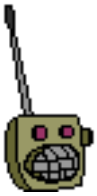

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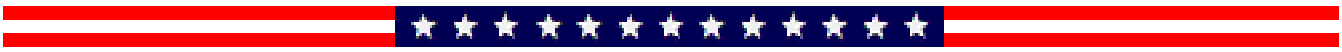

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AC6V E-MAIL ADDRESS

This book is six months in the writing and features the advice of several noted Amateurs and technical gurus.

It is intended for those entering the world of FM Repeaters and Amateur Radio. The guide is written for new users with or without a technical

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FM 101x is comb-bound so the latest updates will be inserted prior to all orders as we print on demand. This way all updated URL's, reader feedback, and corrections will be in your manual. As further addendums are received, they will be posted here under
MANUAL UPDATES

SEE E-HAM REVIEWS

TESTIMONIALS (UNSOLICITED)

Universal Radio comments: Many new hams find repeater operation intimidating. This book lets you skip that awkward early stage. It provides the knowledge and confidence you need to enjoy repeater operation immediately.

I would like to send a thanx to you for your book on repeaters. Just starting out in the hobby. I could not have asked for a better guide to the world of fm repeaters. I have now started working 2 m ssb and have logged over 17 grid squares in 4 months. Thanx again for a great start in the hobby. 73 George KG6TVC, San Leandro, CA

KF6HBJ writes -- I purchased the FM101x Guide last week and have gone through the material relating to my HT. Basically I purchased it to bring me up to date as I had dropped off from using my radio quite awhile ago. Being a computer guy, I found the material easy to follow and found the essentials pointing me to further readings on web links for increased knowledge and detail. From the first page to the last page, I never felt slighted because I wasn't a seasoned Ham Radio

Operator. Rod's vernacular clearly expresses what I needed to get accustomed to without the academia sort of information that I have read in other technical types of volumes. Thanks Dave KF6HBJ, Temecula CA

Excellent Book. I recently taught a one day ham class for new technician hams. I purchased a copy of the FM101X book for each student. All the students had many compliments for the contents of the books. One comment "including the book in the course was a very good idea" Another comment"FM101X answered all my questions about repeater operation as a new ham - great book"
The new hams loved it! For new hams this is the book! Dick Decker K6SUU, Turlock CA

N5ACM comments on eHam reviews: Get up to Speed Quickly with this Guide. This book is well worth the price. You get current, practical information on topics of interest to both new and old HAMs. Information is presented in a straightforward, succinct and readable manner, the illustrations are basic, but adequate. You don't need or pay for glossy fluff here- the book provides a well thought out treatise on FM ops in one place, from simplex, repeaters, digital and the vagaries of Echolink. Lots of links for further information too. A great place to get up to speed quickly. New HAMS will feel completely comfortable on VHF/UHF after perusing this book. Well Done. N5ACM, Arlington TX

AB7NI Wrote: Rod, Received the book today. I read it, and was impressed. Great minds must think alike. Your comments were EXACTLY what I have been telling new hams for years. Very much interested in buying in volume for the next class I teach. Meanwhile this copy is getting some yellow highlighted sections. Great job! Thanks -- Gary AB7NI Colbert, WA

KG6JPX Wrote -- Bravo on AC6V'S FM101x. Also, the internet website references are great. However I think it would help a lot if you repeated them in the back of the book, grouped according to their purpose. That would help locate them later on when the reader has forgotten the context, chapter and page number where it was first seen.

Author note -- this will be added --see MANUAL UPDATES

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Chapter 1 Introduction - Click Here For Sample

Covers an introduction to the book as well as an example of the bewildering jargon that one will encounter on Amateur Radio Repeaters and where in the book to decipher the alphabet soup and jargon. CTCSS, PL, Offset, Desense, Alligator, Machine, Picket-Fencing, Quieting, Q-Signals, beeps, hang-time, protocol, OM, capturing, etc.

Chapter 2. Your First FM Radio - Click Here For Sample

Information on what to select as your first FM radio. Discusses advantages of Handi-Talkies vs Mobiles, antennas, batteries, features to consider. All mode radios and multi-banders are discussed. Mobile installation, Mobile Antennas, DC power supplies, Noise Abatement, Mobile Power considerations, Coax Considerations, and VSWR checks.

Chapter 3. Operating Simplex - Click Here For Sample

Discusses how to operate simplex (without a repeater). Range to be expected, antennas, cross polarization, and protocol. Includes a complete list of simplex frequencies from 2 meters thru 1.2 GHz including the recommended National Calling frequencies. VHF/UHF DXing techniques.

Chapter 4. How Repeaters Works - Click Here For Sample

Simplified pictorial of a repeater. Covers simplex, half duplex, full duplex operation. Explains offsets, splits, input and output frequencies. Lists the standard offsets for USA repeaters - 10 meters thru 2.4 GHz. Thoroughly covered are CTCSS, PL, SUBAUDIBLE TONES | TONE SQUELCH | DCS, DTCS |TONE BURST | DTMF | BEEPS & BE-BOPS.

Chapter 5. Programming a Rig - Click Here For Sample

Covers the nightmare of programming a radio. How to make a cheat sheet. Using programming cables. A source for a handy wallet size programming accessory. Generic outline for programming any radio.

Chapter 6. Antennas, Power Sources, VSWR, and DeciBels - Click Here For Sample

Covers rubber duck antennas, better antennas such as ¼ wave, ½ wave, ¾ wave and 5/8 wave, Yagis, and J-Poles. Gives numerous URL's and tips for building your own antennas. The battery paragraphs cover Sealed Lead Acid (SLA),

Lithium, NICAD and NIMH Batteries. Power Supplies, VSWR, and dB are covered in a non-technical approach.

Chapter 7. Using Repeaters - Click Here For Sample

This chapter covers how to find repeaters, repeater guides on the internet, complete rundown on repeater protocol, soliciting a conversation, asking for information, jargon heard, radio checks and signal reports, what to say, breaking in, multiple conversations - rotations, nets, roll calls and demos, autopatching. Also listed is a complete listing of 2 Meter repeater pairs, packet simplex, and voice simplex frequencies.

Chapter 8. Phonetics, Q-Signals, and Callsigns - Click Here For Sample

All about Q-Signals used on repeaters. Recommended Phonetics are covered (Alpha, Bravo, etc). A list of all USA callsign districts and classes of license callsigns are explained.

Chapter 9. Funny Repeater Sounds and Myths - Click Here For Sample

Covers the many sounds heard on a repeater, path noise, distortions, alternator whine, ignition noise, capturing, picket fencing, Morse characters, hum, over and under deviation, kerchunk, desense, motor boating, Station is Calling - But No Beep, Q-Signals, CB talk & ten codes (a no no) etc.

Chapter 10. Cops and Jammers - Click Here For Sample

When and when not to play repeater policeperson. How to deal with jammers and interlopers.

Chapter 11. Inside A Repeater - Click Here For Sample

Detailed block diagrams of the elements of a repeater, receivers, transmitters. Plain language non-technical discussions of duplexers, limiters, discriminators, deviation, bandwidth, channel spacing, modulation index, FM modulation.

Chapter 12. IRLP, ILINK,ECHOLINK, WIRES II™, CrossBand Repeating, Long Range Intertie Systems, - Click Here For Sample

Glossary and Jargon -- 16 pages of terms, slang, repeater speak - Click Here For Sample

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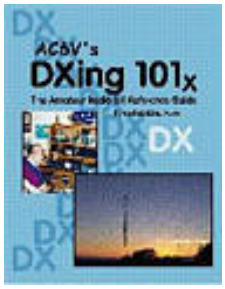
Reviews Summary for DXing 101X			
	Reviews: 19	Average rating: 5.0/5	MSRP: $24.95 plus postage
	Description: This is a great reference book for all DXers.		
	More info: http://ac6v.com/DXSAMPLE.htm		

You can write your own review of the DXing 101X.

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NE5EE	Rating: 5/5	Jul 3, 2004 20:47	Send this review to a friend
An extensive compendium			**Time owned: 0 to 3 months**
I had picked up a copy of FM101x and was quite impressed (I had planned to use it for a followup class for new hams). So I took a look at DX101x too. Wow, it is great! So much info. And lots of stuff I didn't know even tho I have over 200 countries. This book is definitely worth it, right up there with The COmplete DXer.			

GW0ANA	Rating: 5/5	Apr 22, 2004 06:18	Send this review to a friend
A great book very well researched & packed with DX secrets			**Time owned: 0 to 3 months**

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I found it very educational on the subject of HFDX-ing. In fact is was like having your own Elmer in the shack. I have built up a good libary of books over 25 years of "hamming" & I can recomend AC6V`s book DXing 101X as the Definitive guide. Each chapter is very well laid out in a easy to read format. His writing style is great as it is not bogged down with waffle & padding. Its the facts the real facts & nothing but the facts. As I told Rod, I am sure with the advice found in his book I an sure it will be my "Passport to DXCC Honour Roll" 73 de Glyn GW0ANA

VE9VIC	Rating: 5/5	Feb 16, 2004 20:31	Send this review to a friend
a must			**Time owned: more than 12 months**

I HAVE BEEN DXING FOR THE PAst 1o years and
this book is really good for the newcomers
and for the more experimented dxer's.
nice job,it will sure help a lots of news
amateurs specialy if they don't have or
find a elmer to help them,I also have the
complete dxer by w9kni AND LOW BAND dxing
by ON4UN ,all are a must if you want to
improve your skill and have fun,73 best dx.rino

N5VWN	Rating: 5/5	Dec 31, 2003 06:14	Send this review to a friend
It's like, DXing for Dummies			**Time owned: 6 to 12 months**

Best book for DXing for newbies or perhaps anyone. I'm sure even the seasoned pro could glean stuff from here. It's a must have for people like me who know nothing. Well I know something now. This book is layed out logically and is an easy read. I even chuckled a few times. His sense of humor it perfectly timed. But mostly it's so handy and I keep refering too it daily while DXing. It's got usefull tricks to DXing and tips on equipment and procedures. Really enjoyed his DXing secrets. IMHO, probably the most important thing is the frequency listings and their uses. I'm hoping that will help standardize things as far as where the DX is. But I'm amazed how knowledgable AC6V is on this subject. I've got tabs pasted all over it for quick refrence and it is right next to me in the ham shack. . . usually opened to something like frequency list or the "Q" codes. Tons of stuff in there like that. I'm VERY glad I bought this one and I don't buy many "how to" books.

KU4BP	Rating: 5/5	Jul 31, 2003 00:49	Send this review to a friend
Great Book			**Time owned: 6 to 12 months**
This book is a must for new DXers and DX veterans alike. The chapter on DX secrets is an invaluable resource because nobody knows every "trick of trade". Ed KU4BP			

KE6NHS	Rating: 5/5	Jul 26, 2003 15:56	Send this review to a friend
Excellent			**Time owned: 0 to 3 months**
If you are new to HF this book is your new best friend. All of the basics are well covered. Once you get past the basics, you will find a lot of advanced information that can only be acquired through years of experience. After spending 2 weeks with the book I am still finding valuable information and I am constantly referring to the many operating aids provided. Thank you Rod!			

AD6WL	Rating: 5/5	Jun 25, 2003 17:00	Send this review to a friend
Great book			**Time owned: more than 12 months**
This book dosn't have anything it in that I havn't read before, but it has it all in one place. I have read many books, and websites on HF and DXing. This book will save you lots of time by having it all in one easy reference. It is a great book for newcomers to DX and the oldtimers alike.			

N6SDK	Rating: 5/5	Dec 27, 2002 16:21	Send this review to a friend
Great HF'ing Book			**Time owned: more than 12 months**
[redacted]			

This book has helped me time and time again. It is packed with years of useful information, which the author, Rod AC6V, has learned in through his experiences. It is organized in a straight forward easy to understand manner. I use this book whenever I have unanswered questions or simply want to read about High Frequency Ham Radio operation. I highly recommend it!

AG5T	Rating: 5/5	Dec 26, 2002 12:11	Send this review to a friend
Fantastic dx book			**Time owned: more than 12 months**

What a great dx book! When I was called to jury duty I selected THIS BOOK to read while waiting in the jury assembly room! It's a great way to learn and review dxing techniques. Those dxing secrets are great (but don't tell anyone!). I know a lot of work went into this book and I thank and respect all those who worked on it. Another thing that makes it great is the humor mixed in - in fact, I'm asking my local radio club if they want to go on a dxpedition to Charcoal Island in the Barbecues IOTA group! Thank you Rod and everyone for a great book.

KB1GZ	Rating: 5/5	Sep 5, 2002 13:01	Send this review to a friend
Terrific			**Time owned: 6 to 12 months**

Although licensed for several years, been away from Amateur Radio for about 10 years, so getting restarted with all the new technology and information available was a chore, but DX101X made things a whole lot more enjoyable.

The bottom line is this, I have read the book three times and extracted only a small percentage of the available information; and, if you are either new to our hobby or think you have a strong enough ego not to be humbled but to absorb the information contained in this book, you are in for a very pleasant surprise!

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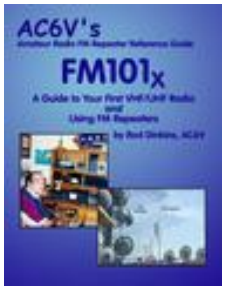
FM101x By AC6V - A Beginners Guide To Using FM Repeaters

Reviews Summary for FM101x By AC6V - A Beginners Guide To Using FM Repeaters			
	Reviews: 6	Average rating: 4.7/5	MSRP: $12.95
	Description: For the new Ham entering the world of FM repeaters and VHF/UHF. Selecting your first rig, Repeater Operations, Finding and Using Repeaters, Antennas, Installation, Power Sources and Batteries, Protocol, Q-Signals, Jargon revealed - 16 page Glossary. PL, Beeps, Timers, Offset, Hang Time, DCS, DTCS |TONE BURST | DTMF etc. Desense, Capturing, Path Noise, Quieting, How FM works.		
	More info: http://ac6v.com/FM101.htm		

You can write your own review of the FM101x By AC6V - A Beginners Guide To Using FM Repeaters.

KC6TRW	Rating: 5/5	Aug 28, 2004 16:29	Send this review to a friend
Great book Rod ! Easy reading and good for the beginner			**Time owned: 3 to 6 months**
Most folks at the Aircraft factory are now asking for this book so they can conteract being mic shy. Good work Rod! Denis Despins KC6TRW			

NE5EE	Rating: 5/5	May 25, 2004 16:20	Send this review to a friend

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A terrific fast start	**Time owned: 0 to 3 months**
A new Tech enters the hobby with an HT and little else. This is the place to pick up the rest of what he or she needs to know. I do a lot of coaching and I will be recommending this book from now on. Two years of experience packed into one weekend of easy, pleasurable reading.	

KF6HBJ	Rating: 5/5	Apr 28, 2004 16:29	Send this review to a friend
Book clearly written to be read and used.			**Time owned: 0 to 3 months**
"I purchased the FM101x Guide last week and have gone through the material relating to my HT. Basically I purchased it to bring me up to date as I had dropped off from using my radio quite awhile ago. Being a computer guy, I found the material easy to follow and found the essentials pointing me to further readings on web links for increased knowledge and detail. From the first page to the last page, I never felt slighted because I wasn't a seasoned Ham Radio Operator. Rod's vernacular clearly expresses what I needed to get accustomed to without the academia sort of information that I have read in other technical types of volumes. Thanks Rod! 73 Dave - KF6HBJ"			

K6SUU	Rating: 5/5	Mar 10, 2004 20:39	Send this review to a friend
excellent book!			**Time owned: 0 to 3 months**
I recently taught a one day ham class for new technician hams. I purchased a copy of the FM101X book for each student. All the students had many compliments for the contents of the books. One comment "including the book in the course was a very good idea" Another comment"FM101X answered all my questions about repeater operation as a new ham-great book" For new hams this is the book! Dick Decker K6SUU			

N5ACM	Rating: 5/5	Apr 30, 2003 14:59	Send this review to a friend

Get up to Speed Quickly with this Guide	**Time owned: 0 to 3 months**
This book is well worth the price. You get current, practical information on topics of interest to both new and old HAMs. Information is presented in a straightforward, succint and readable manner, the illustrations are basic, but adequate. You don't need or pay for glossy fluff here- the book provides a well thought out treatise on FM ops in one place, from simplex, repeaters, digital and the vagaries of Echolink. Lots of links for further information too. A great place to get up to speed quickly. New HAMS will feel completely comfortable on VHF/UHF after perusing this book. Well Done.	

NA4M	**Rating: 3/5**	**Apr 10, 2003 16:37**	**Send this review to a friend**
New Book on FM			**Time owned: 0 to 3 months**
New book on FM repeaters. This will be removed when first review posted.			

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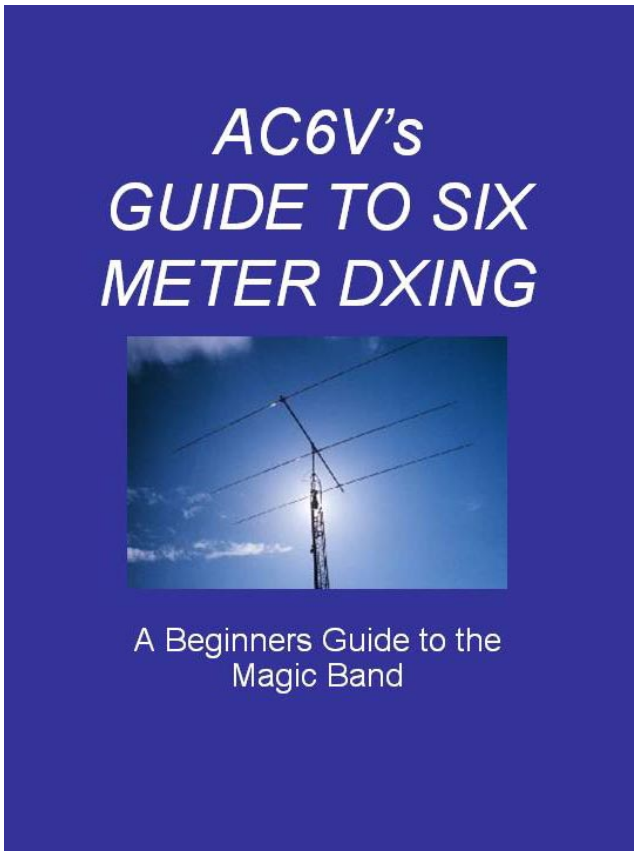

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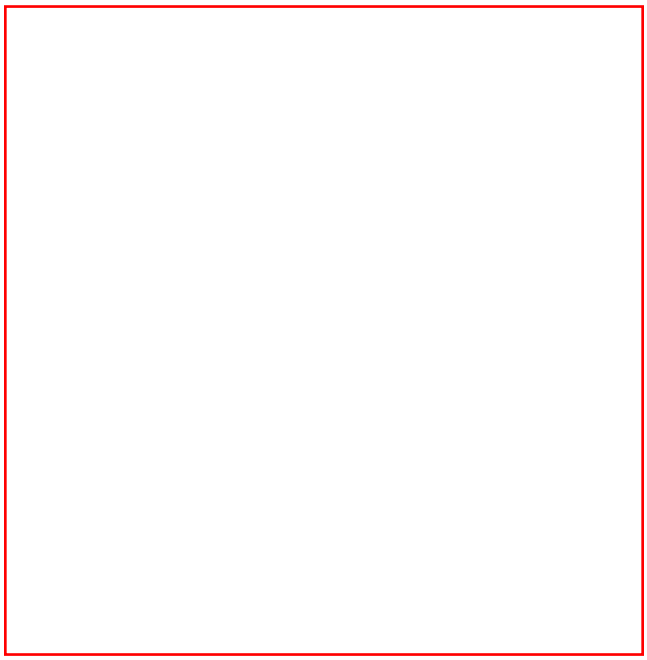

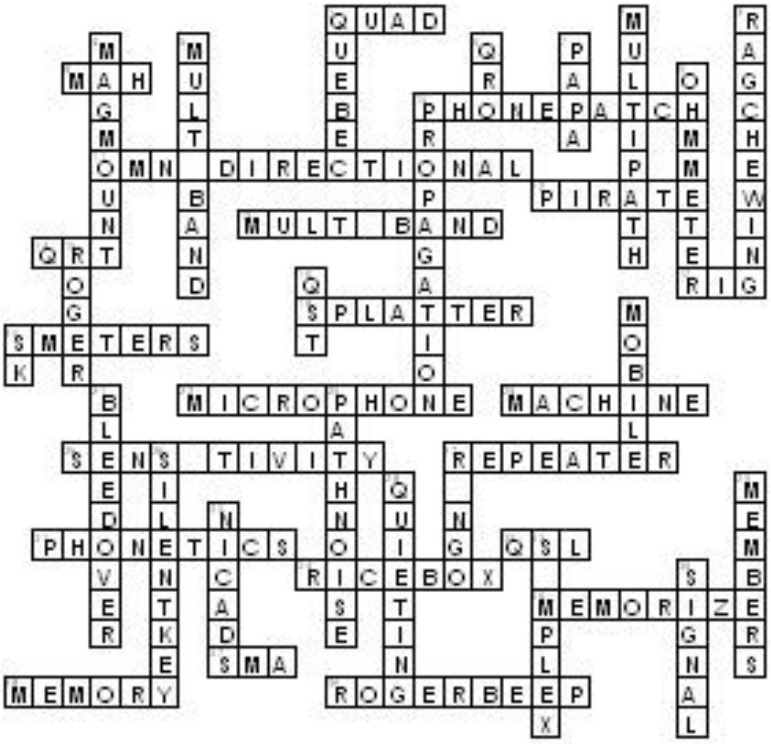

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- Antenna Specialists -- Mobile Antennas -- Via Universal Radio
- .htm
- Anti-oxidation products -- NOALOX, PENETROX, more -- Also See AES Ham Store See On-Line Catalog #14 Page 131-137 or Local Hardware Store -- Also see Coax-Seal
- Anntron Antennas -- 80 thru 10M Vertical, Hamwhip Single Banders, 10 - 11 Meter Base, VHF/UHF
- AN Wireless -- Self Supporting Towers
- Arrow Antennas -- VHF/UHF Antennas includes Satellite Antennas, Yagis, J-Poles, Corner Reflectors
- ARS antenna rotator control with a computer From EA4TX
- Atoc Iron Horse HF Mobile -- Via Universal Radio
- Autek - Antenna Analyzers

- BandHopper -- Mobile/portable/fixed antennas
- Barker & Williamson - Antennas, Coils, Accessories for HF Radio
- Belden Cable -- go to Products and Services, then Belden Catalog -- in the search box -- type in cable number or 50 (75) ohm coax
- Beverage Antenna Transformers - From K1FZ
- BlueSky Masts -- Portable Antenna Masts
  -- Up to 20 meters
- Bozak Antennas - VHf & UHF Antennas
- BuddiPole -- W3FF Products
- Butternut Antennas - Via Bencher Pages
- BUX CommCo ™® Antennas, Connectors, Parts and Accessories.

- C3I® Antennas - VHF/UHF antennas
- Cable Experts
- Cable To Go
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- Decibel Products -- Wide Range Of Antennas
- Degen Designs -- MiniQuad three band, portable, high performance mini-quad antenna
- Delta X-Ray Antennas - - HF Delta Loops and VHF/UHF ground plane antennas
- Diamond Antennas - via RF Parts Company
- N5FC's Notes on the DK9SQ Telescoping Mast
- DWM Communications -- Yo Yo Wind up Dipole Antenna, Tiny Tenna, Travel Tenna
- DX Engineering - Boom-to-Mast Plates, Element-to-Boom Plates, Element Clamps, and Harbach Saddle Clamps.
- DX System Radio - HF thru UHF Antennas

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- Ventanna HFP-2 10 Foot HF Antenna -- 6.5 to 30 MHz
- Force 12 Antennas

- GB Antennas - From The Netherlands
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- GeoTool - Truck Antenna Mount
- Giovannini Antennas - Italian Site -- D2T Antenna 1.5/200 MHz, 1 KW, Low band DXing Antennas, Broad Band Antennas 1.8/30 MHz 1 KW, Quads, Fiberglass Masts, Rotators
- Glen Martin Engineering - Towers, Hazer, Telescoping Tower Trailer, Roof-Top Towers, Kevlar Guy Cable & Rope
- GLA Antennas - Texas BugCatcher
- Guided Wave Technology -- analyze and design rectangular waveguide filters.

- HamAntenna -- Canadian Mfg
- HamStick Antennas
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- Heights Tower Systems
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- Hybrid Quad Antenna -- From TGM Communications
- Hy-Gain - By Telex -- New page - Diect Marketing
- HyperLink Technologies - Antennas, Cable, Accessories
- Hy Power Antenna Company - Several wire antenna designs
- ---
- IIX Equipment Ltd ---- Via W9IIX -- GinPoles, Rotor Mounts, Lots of Tower Accessories

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- K1CRA Radio Web Store -- RF Connectors, Liquid Tape, Plasti-Dip Products
- K3IWK -- VHF & UHF Antennas
- KB6KQ Loop Antennas -- 10 M through 70 cM Stackable Loops
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- M2 Antennas
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- Max Gain Systems - RF coaxial switches, Gin Poles, Antenna Accessories, HF quads
- MaxRad Antennas -- VHF and UHF Antennas
- MFJ catalog -- Antennas and Antenna Analyzers
- MoonRaker Antennas -- UK Manufacturer
- Mosley HF Beam & Vertical Antennas Mosely Home page
- ---

- N1LO's Tower Page -- HomeBrew Tower, Guying, Safety, Etc
- Nemal Electronics International -- Connectors and Cables
- Nil-Jon VHF/UHF Base -- Via Universal Radio
- NI8F HF Dipole Antennas -- Via Universal radio
- NOALOX -- Anti-Oxidation Product Also See AES Ham Store See On-Line Catalog #14 Page 131-137 or Local Hardware Store
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- Smiley Antennas
- Solarcon A99 - Via Universal Radio
- Solarcon IMAX 2000 Base Antenna -- a 5/8 version of the A99
- Solder-It -- All Kinds Of Solder Products
- Sommer Antennas
- Spider Antennas -- From Multi-Band Antennas
- SpiRo's MultiBand Antennas
- SSB Electronics -- Big Wheel Antennas 2M & 70cm
- Stealth Antennas -- HF Mobile Antennas By VE7BOC
- Steppir Antennas -- Fluid Motion
- Stuf -- Dielectric Water-Proof Connector Filler - See AES Ham Store See #14 Page 131-137 or Local Hardware Store
- Super Antennas - Vern Wright's HF Mobile Antennas, MP1 etc.

- T-Hunting Equipment
- TarHeel Mobile Antennas
- Telewave -- Combiners, MultiCouplers, cavities
- Telex -- Hy-Gain Products
- Tennadyne Corp -- Log Periodics
- Texas BugCatcher And Texas Twister- GLA Antennas
- Texas Towers
- T.G.M. Communications -- Mini-Quads
- Thrust Bearing FAQ -- From The YCCC
- Times Microwave Systems -- Coax and Connectors
- Titanex Antennas
- Top Ten Devices, Inc - Data decoders and antenna switching
- Traffie Technologies -- Hex Beams
- Transel Technologies. -- Antenna Mounts, Thread Mounts, HF and VHF/UHF Antennas
- Trilogy Communications -- High Grade Cable
- Trival Antennas -- From Slovenia

- Unadilla Antenna Mfg Co. -- Baluns, Traps, Coaxial Relays, Insulators, Wire
- Universal Radio -- Large Selection of Antennas, Towers, Rotors, Tuners, Switches, Coax
- US Tower

- Valcom -- Fibreglass Whip Antennas
- Valor VHF/UHF Antennas -- Via Universal Radio

- Vectronics -- Tuners, Filters, Dummy Loads, Keyers, Wattmeters
- VENTENNA -- A VHF/UHF vent pipe antenna -- Click Here For eHam Reviews
- Ventanna HFP-2 10 Foot HF Antenna -- 6.5 to 30 MHz

- W3FF Antennas -- Buddipole™, the new Buddistick™ for HF Portable operations.
- W7FG - Ladder Line and Antennas. 600 Ohms. Doublets and G5RV antennas
- WB0W, Inc - Antennas, Towers, Hazers, more
- Wellbrook Communications -- Antenna Loops for MEDIUM WAVE, LONG WAVE & SHORT WAVE

- WideBand Loop Antenna -- From Maarten Haag
- Wilderness Radio - For the outdoor QRP enthusiast
- Wilson Antennas
- WireMan Inc -- Wires, Coax, Baluns A Bunch
- WorldRadio -- SD-20 Antenna Support extends from 47 inches to 20 feet. It weighs less than 2 pounds. Go to Products.
- W5AJ Tower Raising
- WX0B -- StackMatch - for stacking, phasing, and switching antennas

---

- XXTowers -- Experts in amateur radio installations of all types and sizes.

Antenna Tuners

Alinco Electronics
Alpha Delta
Ameritron
Elecraft
ICOM
Kenwood
LDG Electronics
MFJ Enterprises, Inc.
PalStar
SGC, Inc.
Ten-Tec
Vectronics
Yaesu

- # Mobile Antenna Manufacturers
-
- **FAQ ON GLASS MOUNT ANTENNAS & GENERAL INSTALLATION INFO**
-
- MOBILE ANTENNA SHOOTOUT TEST RESULTS
- AntenneX
- Antenna Specialists
- BandHopper
- Carolina Cyclone Mobile Antennas - From K4KWH
- Comet/NCG
- Diamond MV3A Mobile Antenna System
- DK3 Screwdriver - N7LYY
- GS Manufacturing Mobile Antennas
- HamStick Antennas
- Hi-Q Antennas
- High Sierra Antennas
- Hustler Antennas
- K4POZ Mobile Antennas
- Lakeview - Hamsticks
- Larsen Antennas
- New-Tronics (Hustler)
- RadioWare
- The ScrewDriver.Com -- High quality full legal limit with mount and digital controller at an affordable price.
- Spider Antennas
- Super Antennas - W6MMA
- Tarheal Antennas
- Terlin Outbacker
- Texas BugCatcher

COMMERCIAL ANTENNAS USED FOR STEALTH

ABOUT CC&R's AND RESTRICTIONS

BuddiPole -- W3FF Products

B&W Balcony/Window Antenna - AP-10A

CliffDweller II antenna
End-Fedz -- full length half wave dipoles, but with an important difference. The coax connector is at one end of the dipole, where it is most needed. From Universal Radio
EZ HANG -- SlingShot & Reel Device for installing wire antennas
Force 12- 9 Foot Vertical
Force 12 Aluminum Flag Poles & Antennas
HamSticks
High Sierra Antennas - Mobile HF antennas, New El Dorado for Restricted Locations
Isotron Antennas
MFJ Apartment Antenna 40 Meters to 2 Meters MFJ-1622
MFJ Super Loop Antennas
Outbacker Antennas -- Terlin
Quicksilver Radio Products MinuteMan™ HF Portable Antennas
The ScrewDriver.Com -- High quality full legal limit with mount and digital controller at an affordable price.
Stealth Antennas -- HF Mobile Antennas By VE7BOC
Texas BugCatcher - GLA Antennas
Texas Twister - GLA Antennas
Ventennas -- The Forbes Group
Ventanna HFP-2 10 Foot HF Antenna -- 6.5 to 30 MHz !!!
Vern Wright's HF Mobile Antennas, MP1
WeatherVane Antennas

Selected Antenna Books

It Is Back In Print -- Ed Knoll's "Easy Up Antennas" - A good One

New From Ed Knoll "Dipole and Long Wire Antennas"

The Arrl Antenna Book
Buy the Book Today!

The Radio Amateur Antenna Handbook By William I. Orr
Buy the Book Today!

Cubical Quad Antennas : How to Build and Adjust Quads -- by William I. Orr, Stuart D. Cowan
Buy the Book Today!

Lew McCoy on Antennas
Buy the Book Today!

Beverage and Longwire Antennas Design and Theory -- by Chuck Hutton
Buy the Book Today!

ON4UN's Low Band DXing -- by John Devoldere -- Great Source For Low band Antennas
Buy the Book Today!

OVER 300 ANTENNA BOOKS FROM AMAZON.COM
SELECT Books then TYPE Antenna in the KEYWORD BOX

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amazon.com

Enter keywords...

HomeBrew Ur Antenna?? -- Visit The AC6V Antenna Homebrew Links

Antenna & Tower Mail Lists -- Subscription Info

Coax Cable Charts With Losses, Power Ratings, etc

Antenna Software -- Visit The AC6V Antenna Software Links

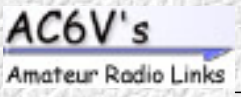
Antennex -- The Ham Antenna Magazine -- all about antennas -- on line

Tower Talk Mail List Reflector -- Subscription Info

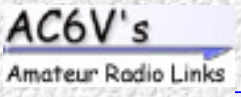
CAN'T FIND THE DEALER OR VENDOR? TRY THESE.

BACK TO THE MASTER DEALER PAGE

- TIS FIND -- From The ARRL



GO BACK TO THE BIG INDEX



GO BACK TO THE QUICK INDEX

AC6V BRAG TAPE

Lsst Update: September 02, 2004

Open links in a new window

SHORTWAVE, LONGWAVE, DC TO DAYLIGHT

NOTE: THE SUBJECT OF SWL IS TOO LARGE TO COVER ON ONE PAGE
THUS PAGES ARE SET UP AS FOLLOWS:

This Page -- see Table Below
Great SWL Frequency Listings
Great SWL Pages and Links
Frequency Band Plans
Spectrum Chart -- See This One

Quick Click

THE GATEWAY TO DC TO DAYLIGHT
Click On Your Topic

AM, FM & TV Stations	Frequency Listings, SWL	QSLing SW Stations
Amateur Radio Listening	Logging Prog - General & Ham	Radio History
Antennas & Grounding	Logging Programs SWL	Receiver Reviews
Bands, SWL	LongWave	Scanners
Books -- SWL & Scanner	Manuals, Radio	Shortwave Web Pages
Books -- AC6V Book Store	Modulation Modes	Software, SWL
Citizens Band	Propagation Software	SWLing The Ham Bands
Dealers & Mfg's, SWL	QSL Forms and QSL Cards	Technical Info For SWLers
FAQ's, SWL		Web Controlled Radios

TO SEARCH THE PAGE
USE YOUR BROWSER FIND COMMAND
OR USE CNTRL F

BOOKS FOR SHORT WAVE LISTENERS

Buy Your Books Thru The AC6V Book Store

RADIOS FOR SHORT WAVE LISTENERS

Buy Your Radios Thru The AC6V Book Store

Easy-Up Antennas For SWL and Hams -- Back In print By Edward Noll -- SEE THIS ONE

Fifty Ways to Improve Your Short Wave Listening From DWM Communications

SCANNER BOOKS

Radios and Reception - Testing and Tips From Philly Talk Radio Online

SWL Receiver Reviews -- From Radio Netherlands

What You Need to Hear Shortwave Stations -- Receivers and Antennas

Short Wave Receiver Reviews -- From eHam.net -- Over 71 Short Wave Receivers

Shortwave Receivers Survey --- Specs, Features and Comprehensive Reviews

Shortwave Receiver Reviews From Strong Signals Resource Page

Shortwave Receiver Reviews -- Via RadioIntel.com

Shortwave Receivers Review -- From Durham Radio

Modern Shortwave Receivers Survey -- From DXing.com

Selecting A Shortwave Receiver -- From DXing.com

Super heterodyne Receiver Theory -- includes Sensitivity, Dynamic Range, more

ICOM PCR 1000 Resource Page -- Test Drive & PCR Frequency Files

Yaesu FRG-7 Information

Sangean ATS-909 Review - From Radio Netherlands

Synchronous Demodulation --- What is it and how it works

SWL SOFTWARE

Propagation Software Is At The Propagation Pages

Radio Explorer -- By Dmitry Nefedov -- Graphic viewer for shortwave schedules (ILGRadio and HFCC), displaying data in a tree-like way and on a Gantt.chart. It also features a world map with greyline and MUF coloring. It is developed in Java and runs on multiple platforms.

Radio Listener's Database integrating the power of the top five broadcast databases available today

LOGGING PROGRAMS

ShortWave Log (SWLog) -- Logging and Radio Control for SWLing

B-Log Version 1.1 -- From AB9B

WRLog v1.10 -- From Paraclete InfoSys

DXtreme™ produces powerful and easy-to-use logging applications for radio enthusiasts such as: Shortwave DXers and Listeners

GenLog --- contests and as a general logger and SWL also.

WLOG2000 LogBook for OM SWL BCL CB -- Database logbook for OM SWL BCL & CB with utility like DXCLuster DXtelnet PSK PSK-Pbbs (TNC or PC sound card) radio rotor CDbook interface and many other features. Free Update at homepage.

SWLTool (Version 1.4.5) ... a utility to read and display ILG radio data files

Many More Ham Logging Programs which you might use for SWL Logging.

SWL ANTENNAS

ALSO SEE SWL DEALERS

Easy-Up Antennas For SWL and Hams -- Back In print By Edward Noll -- SEE THIS ONE

Amandx Antenna Pages
AM Antennas -- by Bruce Carter
AM Loop Antennas
Antenna Advice -- From Radio Netherlands
Antennas A Bunch -- From Hard-Core-DX
Antenna Basics and Theory -- From Ian C. Purdie, VK2TIP
Antenna Basics and Theory -- Excellent Tutorial From Scott's Pages
Antenna Basics and Theory -- Excellent From Integrated Publishing
Antenna Basics and Theory -- From Sub-TV
Antenna Dimension Calculators -- From The Antenna Elmer - Click On Antenna Type
Antenna Dimension Calculators -- Inverted Vee And Dipoles
Antenna Dimensions Calculator Dipoles-- From AMANDX
Antenna Gain -- From L. B. Cebik
Antennas - Low Noise, Grounds for low noise, longwire impedance matching
Antenna Loops -- From Wellbrook Communications
Antennas, SWL -- Wire, Loops, Invisible, Feeding, Grounding, Many More From HCDX

Antennas For SWLs --- From K6QGH
Antenna testing and reviews -- From Philly Talk Radio Online
Antenna Tuner Theory -- From The ARRL
Beverage Antennas
Broomstick Special The Broomstick Special, Part 2
Choosing Wire For An Antenna
Dipole & Inverted Vees Antennas
Doug's Antenna Transformer Page
DWM Communications -- Yo Yo Wind up Dipole Antenna, Tiny Tenna, Travel Tenna
Feed Lines For Antennas -- From HCDX
Frankenstein Monster Antenna -- From Michael's SW and DXing Page
G5RV Antenna For Shortwave -- From Michael's SW and DXing Page
Great White North Antenna
Grounding -- For Safety and For Noise Reduction
Grounding is key to good reception
Height Of Dipoles - Patterns By AA3RL
ICE180 Matching Unit -- From HCDX
Impedance Transformers for Receiving Antennas -- John Bryant
Inverted L Antenna -- From ARNIE CORO (CO2KK)
K9AY Directional Terminated Loop Antenna
Long Wire Antennas -- Via AMANDX
Loop Antennas -- Discussion Group - from Yahoo Forums
Loops for AM BC Band - From Dave's Antenna Loop page
Low Noise Antenna Connection -- Via Boston Area DXers
PAR High-Performance Shortwave Antenna HF End Fed SWL antenna
Random Wire Antenna and Improved Random Wire Antenna -- From CW's Radio Pages
T2FD Terminated Tilted Folded Dipole
Tuned Loop AM Broadcast Antenna -- From VK2ZAY
WellBrook Loop Antennas
WideBand Loop Antenna -- From Maarten Haag

Commercial SWL Antennas I -- From Universal Radio

Commercial SWL Antenna II -- From Alpha Delta

SWL FORUM -- FROM YAHOO

DXing.Com -- All about SWLing and DXing -- From Universal Radio

FAQ'S ABOUT SHORTWAVE RADIO --From National Association of Shortwave Broadcasters, Inc.

Shortwave Radio Information -- for Beginners by Lou Castino

SWLing The Ham Bands

Digital Signals FAQ -- Loaded -- From WUN

Radio H.F. Catalogue & Newsletter -- By Sheldon Harvey

Short Wave Amateur Radio Listening

MODULATION AND MODES

How Modulation Modes Work -- Very Visual Explanations for AM, FM, SSB, DSB, QAM

Modulation Codes --- AJE, A1A, Etc

Modes & Modulation -- CW, AM, SSB, USB, LSB, FM, FSK, AMTOR, FEC, PACKET, Spread Spectrum -- From DXing.com

All About SSB, DSB, ECSS, more -- History Of SSB By DJ4BR Peter Weber

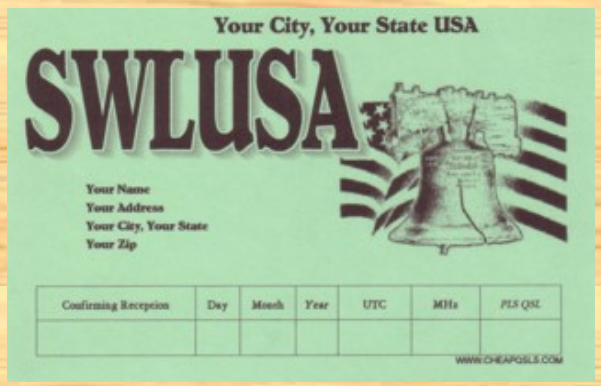
Modulation Schemes --- AM, FM, Pulse Technical Discussions

ALSO SEE OPERATING MODES -- AMTOR, APRS, ATV, CLOVER, CW, Digital, EME, SETI, Astronomy, FAX, GMRS, GPS, GTOR, Laser, Low Freq, Meteor Scatter, Microwave, NAVTEX, PACKET, PACTOR, PSK31, QRP, RTTY, Satellites, Spread Spectrum, SSTV, T-Hunts, WEFAX

QSLING SW STATIONS, SAMPLE QSL FORMS AND CARDS, QSL

GALLERIES

SWL QSL CARDS FROM CHEAP QSL's.COM



All About Reporting and QSLing -- Via DXing.com

Reporting and QSL - Via Amandx

A realistic guide to reception report writing in the new millennium -- From Radio Netherlands

Reporting Codes --- SIO and SINPO -- From Radio Netherlands

QSL Form --- From China Radio International/Shortwave

QSL Form --- From K6QGH

SWL QSL Card Museum -- Wow -- Loaded A massive collection of SWL QSL cards from around the world

Get Ur Own SWL Callsign

GENERAL FREQUENCY LISTINGS

FREQUENCY BAND PLANS -- FOR ALL KINDS OF SERVICES Click Here

SWL FREQUENCY LISTINGS

BAND	BAND

120 m ___2300 -- 2495 kHz	25 m ____11.500 -- 12.160 MHz
90 m ____3200 -- 3400 kHz	22 m ____13.570 -- 13.870 MHz
75 m ____3900 -- 4000 kHz	19 m ____15.030 -- 15.800 MHz
60 m ____4750 -- 5060 kHz	17m ____ 17.480 -- 17.900 MHz
49 m ____5730 -- 6295 kHz	16m ____ 18.900 -- 19.020 MHz
41m ____ 6890 -- 6990 kHz	13 m ____21.450 -- 21.750 MHz
41 m ____7100 -- 7600 kHz	11 m ____25.670 -- 26.100 MHz
31 m ____9250 -- 9990 kHz	

The "Radio Regulations" of the International Telecommunications Union define the shortwave broadcast bands. At the 1992 World Administrative Radio Conference, new bands were created and existing bands were expanded. The band limits in the following table reflect the WARC-92 agreements and broadcast band expansions used on a non-interfering basis (e.g., the 41 m band starting at 6890 kHz on a non-interference basis): Although the allocations do not become official until 2007, in practice many stations have already started using the expanded portions under the motto "use it or lose it." Additionally, there are a few stations that broadcast outside the band edges above, e.g., Iran on 9022 and a number of African stations around 9200 kHz. Clearly, receivers with continuous coverage between 1.6 and 30 MHz are preferred.

FREQUENCY LISTINGS WORLD WIDE -- HUNDREDS - Click Here

GREAT SWL FREQUENCY LISTINGS

FROM MONITORING TIMES

Listner Laws

Airshow Frequency List

Monitoring NASA Communications (Space Shuttle et al)

Hot 1000 HF Frequencies

Monitoring Times Frequency Exchange

Monitoring Times Frequency Reference Library

Comprehensive Shortwave Broadcasting Schedule By Eike Bierwirth, Leipzig, Germany

ID, Interval signals, signature tunes, and identification announcements from international, domestic, and clandestine radio stations around the world

SWL Stations that are scheduled to broadcast in English at the moment you request the page.

SWL Stations -- enter a frequency, and see all those who are scheduled to use that frequency.

SWL Stations -- By Time Or Country -- From PrimeTime ShortWave

Optimal Frequencies By Time For This Month

Hot 100 SWL Stations -- From ICOM America

International Broadcasters on Short-wave Radio -- From Steve R. Adams

Top 1000 HF SWL Frequencies -- From Monitoring Times

Hundreds Of SWL Stations Transmitting In English to The USA

NASWA Shortwave Loggings -- Loaded

Prime Time Short Wave -- Your guide for English shortwave broadcast schedules

Also See Great SWL Web Pages

AM, FM & TV STATIONS

50KW AM Night Stations -- USA With Canada and Mexico Listings

Finding AM/FM Stations Near You (USA & World Wide)

FCCINFO Search -- Find AM, FM, TV Stations in the USA

FCC Data Base - Several search options available

Medium Wave Radio -- From Radio Netherlands

AM Broadcast Station Locator -- From The FCC

FM Broadcast Station Locator -- From The FCC

TV Broadcast Station Locator -- From The FCC

TV DXing -- From Jeff Kadet

Travelers AM Locator -- From The FCC

FM and NTSC TV Propagation Curves Calculation -- From The FCC

For Broadcast Professionals -- Quick Access to Related Links. -- From Radio 411

U.S. AM Radio Stations Radio-Locator -- (Formerly known as "The MIT List of Radio Stations on the Internet")

AM or FM Data Base -- From Elliott Broadcast Services

TvRadioWorld -- Internet Broadcast Directory and Radio/Television Stations Web Listings

Radio History -- 16 Top Pages Of Radio History

SCANNERS

BOOKS FOR SCANNER ENTHUSIASTS

Scanners & Secret Frequencies (Electronic Underground Series, Vol 3) by Henry L. Eisenson
Buy the Book Today!

SCANNER WEB PAGES & FREQUENCIES

FROM MONITORING TIMES

Listner Laws

Airshow Frequency List

Monitoring NASA Communications (Space Shuttle et al)

Hot 1000 HF Frequencies

Monitoring Times Frequency Exchange

Monitoring Times Frequency Reference Library

FCC Data Base - Several search options available

-- Check Em Out

Introduction to Scanning -- by Bob Parnass, AJ9S -- If you own a Scanner -- these are a MUST READ

PerCon Data Finder -- Exhaustive online source of frequencies

Police Calls -- Visit Radio Shack for "Police Call" series of books

Selecting A Scanner -- From DXing.com

Scanner Users FAQ -- From Rich Wells N2MCA

Scanners Reference Guide -- By Clay Irving N2VKG

10-Codes -- Police, Fire, etc Usage

GO BACK TO THE BIG INDEX

GO BACK TO THE QUICK INDEX

AC6V BRAG TAPE

This Page Last Updated:September 03, 2004

Check Here If You Want To Open Links In A New Window

Antenna Software

Compiled By AC6V

NOTICE:

The following is a compilation of URLs found on search engines and news groups. The Author makes no endorsement nor has any pecuniary interest in any of the listings.

CAVEAT EMPTOR applies as in all things.

Antenna Design Freeware -- 1D and 2D arrays from MEI Software

Antenna Designing Software

Antenna Feed System Modeling Software by Dan Maguire, AC6LA -- includes Smith chart tutorial

Antenna Shareware A Bunch -- See Goldmine Software one thru five -- not to be missed!!

Antenna Software A Ton -- From The Spread Spectrum Scene -- Not to be missed!!

Antena Solver -by Dr. David Fluckiger, KJ5AT

Antenna Software

Antenna Design and Software From G4FGQ -- Loaded -- many antenna design programs, coax rating, ferrites and toroids, Groundwave propagation, much much more.

Antenna Software -- Loaded -- Via OK1RR

Antenna Software by W7EL -- EZNEC, EZNEC pro, ELNEC

Antenna Software From QRZ

Cubical Quad Antenna Calculator -- From KD6DKS

EMMCAP Software for modeling of arbitrarily shaped wire structures and the computation of their electromagnetic behavior, including radiation and scattering problems.

EZNEC 3.0 -- For for Windows 95/98/NT/200

Grating Solver Development Company - Antenna Solver V1.1a and GSOLVER© V4.20a Highly recommended by a local antenna guru.

Mac Antenna Master -- From Black Cat Systems

The MININEC Professional Series

MMANA (MM Antenna Analyzer) BY JE3HHT - Makoto Mori

MultiNEC 2.0 -- -- Automate and animate your antenna modeling and propagation predictions -- From AC6LA

NEC4WIN® -- Antenna Simulation Software for Windows © NecView® Windows© interface for NEC2 From VE2GMI Madjid Boukri

Nittany Scientific, Inc. -- Antenna Analysis with the User in Mind

PolarPlot -- Polar diagram of your rotatable beam antenna actually looks like where it is operating. From Bob, G4HFQ

Power Loss/ dB Calculator _by W4SM

RF Calculator Page From WA7CS -- Over 40 Programs to help with RF design and antennas

YagiStress -- The Ultimate Mechanical Design Software For Antenna Designers. -- By Kurt Andress, NI6W

YAGIMAX

YTAD -- by N6BV -- Determine antenna gain based on the slope of terrain

Quick Yagi -- Via WA7RAI's Home Page -- yagi auto-design, auto-optimize, modeling software

---

Commercial Antennas -- Visit AC6V's Antenna Mfgrs Links

HomeBrew Antennas -- Visit AC6V's Antenna Project Links

AC6V's Amateur Radio Links **GO BACK TO THE BIG INDEX**

AC6V's Amateur Radio Links **GO BACK TO THE QUICK INDEX**

AC6V BRAG TAPE

Last Updated: August 21, 2004

NONRESIDENT TRAINING COURSE

SEPTEMBER 1998

Navy Electricity and Electronics Training Series

Module 10—Introduction to Wave Propagation, Transmission Lines, and Antennas

NAVEDTRA 14182

DISTRIBUTION STATEMENT A: Approved for public release; distribution is unlimited.

Although the words "he," "him," and "his" are used sparingly in this course to enhance communication, they are not intended to be gender driven or to affront or discriminate against anyone.

DISTRIBUTION STATEMENT A: Approved for public release; distribution is unlimited.

PREFACE

By enrolling in this self-study course, you have demonstrated a desire to improve yourself and the Navy. Remember, however, this self-study course is only one part of the total Navy training program. Practical experience, schools, selected reading, and your desire to succeed are also necessary to successfully round out a fully meaningful training program.

COURSE OVERVIEW: To introduce the student to the subject of Wave Propagation, Transmission Lines, and Antennas who needs such a background in accomplishing daily work and/or in preparing for further study.

THE COURSE: This self-study course is organized into subject matter areas, each containing learning objectives to help you determine what you should learn along with text and illustrations to help you understand the information. The subject matter reflects day-to-day requirements and experiences of personnel in the rating or skill area. It also reflects guidance provided by Enlisted Community Managers (ECMs) and other senior personnel, technical references, instructions, etc., and either the occupational or naval standards, which are listed in the *Manual of Navy Enlisted Manpower Personnel Classifications and Occupational Standards*, NAVPERS 18068.

THE QUESTIONS: The questions that appear in this course are designed to help you understand the material in the text.

VALUE: In completing this course, you will improve your military and professional knowledge. Importantly, it can also help you study for the Navy-wide advancement in rate examination. If you are studying and discover a reference in the text to another publication for further information, look it up.

1998 Edition Prepared by
FCC(SW) R. Stephen Howard and CWO3 Harvey D. Vaughan

Published by
NAVAL EDUCATION AND TRAINING
PROFESSIONAL DEVELOPMENT
AND TECHNOLOGY CENTER

NAVSUP Logistics Tracking Number
0504-LP-026-8350

Sailor's Creed

"I am a United States Sailor.

I will support and defend the Constitution of the United States of America and I will obey the orders of those appointed over me.

I represent the fighting spirit of the Navy and those who have gone before me to defend freedom and democracy around the world.

I proudly serve my country's Navy combat team with honor, courage and commitment.

I am committed to excellence and the fair treatment of all."

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2. Radio Wave Propagation 2-1
3. Principles of Transmission Lines 3-1
4. Antennas 4-1

APPENDIX

I. Glossary AI-1

INDEX INDEX-1

NAVY ELECTRICITY AND ELECTRONICS TRAINING SERIES

The Navy Electricity and Electronics Training Series (NEETS) was developed for use by personnel in many electrical- and electronic-related Navy ratings. Written by, and with the advice of, senior technicians in these ratings, this series provides beginners with fundamental electrical and electronic concepts through self-study. The presentation of this series is not oriented to any specific rating structure, but is divided into modules containing related information organized into traditional paths of instruction.

The series is designed to give small amounts of information that can be easily digested before advancing further into the more complex material. For a student just becoming acquainted with electricity or electronics, it is highly recommended that the modules be studied in their suggested sequence. While there is a listing of NEETS by module title, the following brief descriptions give a quick overview of how the individual modules flow together.

Module 1, *Introduction to Matter, Energy, and Direct Current,* introduces the course with a short history of electricity and electronics and proceeds into the characteristics of matter, energy, and direct current (dc). It also describes some of the general safety precautions and first-aid procedures that should be common knowledge for a person working in the field of electricity. Related safety hints are located throughout the rest of the series, as well.

Module 2, *Introduction to Alternating Current and Transformers,* is an introduction to alternating current (ac) and transformers, including basic ac theory and fundamentals of electromagnetism, inductance, capacitance, impedance, and transformers.

Module 3, *Introduction to Circuit Protection, Control, and Measurement,* encompasses circuit breakers, fuses, and current limiters used in circuit protection, as well as the theory and use of meters as electrical measuring devices.

Module 4, *Introduction to Electrical Conductors, Wiring Techniques, and Schematic Reading,* presents conductor usage, insulation used as wire covering, splicing, termination of wiring, soldering, and reading electrical wiring diagrams.

Module 5, *Introduction to Generators and Motors,* is an introduction to generators and motors, and covers the uses of ac and dc generators and motors in the conversion of electrical and mechanical energies.

Module 6, *Introduction to Electronic Emission, Tubes, and Power Supplies,* ties the first five modules together in an introduction to vacuum tubes and vacuum-tube power supplies.

Module 7, *Introduction to Solid-State Devices and Power Supplies,* is similar to module 6, but it is in reference to solid-state devices.

Module 8, *Introduction to Amplifiers,* covers amplifiers.

Module 9, *Introduction to Wave-Generation and Wave-Shaping Circuits,* discusses wave generation and wave-shaping circuits.

Module 10, *Introduction to Wave Propagation, Transmission Lines, and Antennas,* presents the characteristics of wave propagation, transmission lines, and antennas.

Module 11, *Microwave Principles,* explains microwave oscillators, amplifiers, and waveguides.

Module 12, *Modulation Principles,* discusses the principles of modulation.

Module 13, *Introduction to Number Systems and Logic Circuits,* presents the fundamental concepts of number systems, Boolean algebra, and logic circuits, all of which pertain to digital computers.

Module 14, *Introduction to Microelectronics,* covers microelectronics technology and miniature and microminiature circuit repair.

Module 15, *Principles of Synchros, Servos, and Gyros,* provides the basic principles, operations, functions, and applications of synchro, servo, and gyro mechanisms.

Module 16, *Introduction to Test Equipment,* is an introduction to some of the more commonly used test equipments and their applications.

Module 17, *Radio-Frequency Communications Principles,* presents the fundamentals of a radio-frequency communications system.

Module 18, *Radar Principles,* covers the fundamentals of a radar system.

Module 19, *The Technician's Handbook,* is a handy reference of commonly used general information, such as electrical and electronic formulas, color coding, and naval supply system data.

Module 20, *Master Glossary,* is the glossary of terms for the series.

Module 21, *Test Methods and Practices,* describes basic test methods and practices.

Module 22, *Introduction to Digital Computers,* is an introduction to digital computers.

Module 23, *Magnetic Recording,* is an introduction to the use and maintenance of magnetic recorders and the concepts of recording on magnetic tape and disks.

Module 24, *Introduction to Fiber Optics,* is an introduction to fiber optics.

Embedded questions are inserted throughout each module, except for modules 19 and 20, which are reference books. If you have any difficulty in answering any of the questions, restudy the applicable section.

Although an attempt has been made to use simple language, various technical words and phrases have necessarily been included. Specific terms are defined in Module 20, *Master Glossary*.

Considerable emphasis has been placed on illustrations to provide a maximum amount of information. In some instances, a knowledge of basic algebra may be required.

Assignments are provided for each module, with the exceptions of Module 19, *The Technician's Handbook*; and Module 20, *Master Glossary*. Course descriptions and ordering information are in NAVEDTRA 12061, *Catalog of Nonresident Training Courses*.

Throughout the text of this course and while using technical manuals associated with the equipment you will be working on, you will find the below notations at the end of some paragraphs. The notations are used to emphasize that safety hazards exist and care must be taken or observed.

WARNING

AN OPERATING PROCEDURE, PRACTICE, OR CONDITION, ETC., WHICH MAY RESULT IN INJURY OR DEATH IF NOT CAREFULLY OBSERVED OR FOLLOWED.

CAUTION

AN OPERATING PROCEDURE, PRACTICE, OR CONDITION, ETC., WHICH MAY RESULT IN DAMAGE TO EQUIPMENT IF NOT CAREFULLY OBSERVED OR FOLLOWED.

NOTE

An operating procedure, practice, or condition, etc., which is essential to emphasize.

INSTRUCTIONS FOR TAKING THE COURSE

ASSIGNMENTS

The text pages that you are to study are listed at the beginning of each assignment. Study these pages carefully before attempting to answer the questions. Pay close attention to tables and illustrations and read the learning objectives. The learning objectives state what you should be able to do after studying the material. Answering the questions correctly helps you accomplish the objectives.

SELECTING YOUR ANSWERS

Read each question carefully, then select the BEST answer. You may refer freely to the text. The answers must be the result of your own work and decisions. You are prohibited from referring to or copying the answers of others and from giving answers to anyone else taking the course.

SUBMITTING YOUR ASSIGNMENTS

To have your assignments graded, you must be enrolled in the course with the Nonresident Training Course Administration Branch at the Naval Education and Training Professional Development and Technology Center (NETPDTC). Following enrollment, there are two ways of having your assignments graded: (1) use the Internet to submit your assignments as you complete them, or (2) send all the assignments at one time by mail to NETPDTC.

Grading on the Internet: Advantages to Internet grading are:

- you may submit your answers as soon as you complete an assignment, and
- you get your results faster; usually by the next working day (approximately 24 hours).

In addition to receiving grade results for each assignment, you will receive course completion confirmation once you have completed all the assignments. To submit your assignment answers via the Internet, go to:

http://courses.cnet.navy.mil

Grading by Mail: When you submit answer sheets by mail, send all of your assignments at one time. Do NOT submit individual answer sheets for grading. Mail all of your assignments in an envelope, which you either provide yourself or obtain from your nearest Educational Services Officer (ESO). Submit answer sheets to:

COMMANDING OFFICER
NETPDTC N331
6490 SAUFLEY FIELD ROAD
PENSACOLA FL 32559-5000

Answer Sheets: All courses include one "scannable" answer sheet for each assignment. These answer sheets are preprinted with your SSN, name, assignment number, and course number. Explanations for completing the answer sheets are on the answer sheet.

Do not use answer sheet reproductions: Use only the original answer sheets that we provide—reproductions will not work with our scanning equipment and cannot be processed.

Follow the instructions for marking your answers on the answer sheet. Be sure that blocks 1, 2, and 3 are filled in correctly. This information is necessary for your course to be properly processed and for you to receive credit for your work.

COMPLETION TIME

Courses must be completed within 12 months from the date of enrollment. This includes time required to resubmit failed assignments.

PASS/FAIL ASSIGNMENT PROCEDURES

If your overall course score is 3.2 or higher, you will pass the course and will not be required to resubmit assignments. Once your assignments have been graded you will receive course completion confirmation.

If you receive less than a 3.2 on any assignment and your overall course score is below 3.2, you will be given the opportunity to resubmit failed assignments. **You may resubmit failed assignments only once.** Internet students will receive notification when they have failed an assignment--they may then resubmit failed assignments on the web site. Internet students may view and print results for failed assignments from the web site. Students who submit by mail will receive a failing result letter and a new answer sheet for resubmission of each failed assignment.

COMPLETION CONFIRMATION

After successfully completing this course, you will receive a letter of completion.

ERRATA

Errata are used to correct minor errors or delete obsolete information in a course. Errata may also be used to provide instructions to the student. If a course has an errata, it will be included as the first page(s) after the front cover. Errata for all courses can be accessed and viewed/downloaded at:

http://www.advancement.cnet.navy.mil

STUDENT FEEDBACK QUESTIONS

We value your suggestions, questions, and criticisms on our courses. If you would like to communicate with us regarding this course, we encourage you, if possible, to use e-mail. If you write or fax, please use a copy of the Student Comment form that follows this page.

For subject matter questions:

E-mail: n315.products@cnet.navy.mil
Phone: Comm: (850) 452-1001, ext. 1728
DSN: 922-1001, ext. 1728
FAX: (850) 452-1370
(Do not fax answer sheets.)
Address: COMMANDING OFFICER
NETPDTC N315
6490 SAUFLEY FIELD ROAD
PENSACOLA FL 32509-5237

For enrollment, shipping, grading, or completion letter questions

E-mail: fleetservices@cnet.navy.mil
Phone: Toll Free: 877-264-8583
Comm: (850) 452-1511/1181/1859
DSN: 922-1511/1181/1859
FAX: (850) 452-1370
(Do not fax answer sheets.)
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NAVAL RESERVE RETIREMENT CREDIT

If you are a member of the Naval Reserve, you will receive retirement points if you are authorized to receive them under current directives governing retirement of Naval Reserve personnel. For Naval Reserve retirement, this course is evaluated at 6 points. (Refer to *Administrative Procedures for Naval Reservists on Inactive Duty*, BUPERSINST 1001.39, for more information about retirement points.)

Student Comments

NEETS Module 10

Course Title: *Introduction to Wave Propagation, Transmission Lines, and Antennas*

NAVEDTRA: 14182 **Date**: _____

We need some information about you:

Rate/Rank and Name: _____ SSN: _____ Command/Unit _____

Street Address: _____ City: _____ State/FPO: _____ Zip _____

Your comments, suggestions, etc.:

Privacy Act Statement: Under authority of Title 5, USC 301, information regarding your military status is requested in processing your comments and in preparing a reply. This information will not be divulged without written authorization to anyone other than those within DOD for official use in determining performance.

NETPDTC 1550/41 (Rev 4-00)

CHAPTER 1

WAVE PROPAGATION

LEARNING OBJECTIVES

Learning objectives are stated at the beginning of each chapter. These learning objectives serve as a preview of the information you are expected to learn in the chapter. The comprehensive check questions are based on the objectives. By successfully completing the NRTC, you indicate that you have met the objectives and have learned the information. The learning objectives are listed below.

Upon completion of this chapter, you should be able to:

1. State what wave motion is, define the terms reflection, refraction, and diffraction, and describe the Doppler effect.
2. State what sound waves are and define a propagating medium.
3. List and define terms as applied to sound waves, such as cycle, frequency, wavelength, and velocity.
4. List the three requirements for sound.
5. Define pitch, intensity, loudness, and quality and their application to sound waves.
6. State the acoustical effects that echoes, reverberation, resonance, and noise have on sound waves.
7. Define light waves and list their characteristics.
8. List the various colors of light and define the terms reflection, refraction, diffusion, and absorption as applied to light waves.
9. State the difference between sound waves and light waves.
10. State the electromagnetic wave theory and list the components of the electromagnetic wave.

INTRODUCTION TO WAVE PROPAGATION

Of the many technical subjects that naval personnel are expected to know, probably the one least susceptible to change is the theory of wave propagation. The basic principles that enable waves to be propagated (transmitted) through space are the same today as they were 70 years ago. One would think, then, that a thorough understanding of these principles is a relatively simple task. For the electrical engineer or the individual with a natural curiosity for the unknown, it is indeed a simple task. Most technicians, however, tend to view wave propagation as something complex and confusing, and would just as soon see this chapter completely disappear from training manuals. This attitude undoubtedly stems from the fact that wave propagation is an invisible force that cannot be detected by the sense of sight or touch. Understanding wave propagation requires the use of the imagination to visualize the associated concepts and how they are used in practical application. This manual was developed to help you visualize

and understand those concepts. Through ample use of illustrations and a step-by-step transition from the simple to the complex, we will help you develop a better understanding of wave propagation. In this chapter, we will discuss propagation theory on an introductory level, without going into the technical details that concern the engineer. However, you must still use thought and imagination to understand the new ideas and concepts as they are presented.

To understand radio wave propagation, you must first learn what wave propagation is and some of the basic physics or properties that affect propagation. Many of these properties are common everyday occurrences, with which you are already familiar.

WHAT IS PROPAGATION?

Early man was quick to recognize the need to communicate beyond the range of the human voice. To satisfy this need, he developed alternate methods of communication, such as hand gestures, beating on a hollow log, and smoke signals. Although these methods were effective, they were still greatly limited in range. Eventually, the range limitations were overcome by the development of courier and postal systems; but there was then a problem of speed. For centuries the time required for the delivery of a message depended on the speed of a horse.

During the latter part of the 19th century, both distance and time limitations were largely overcome. The invention of the telegraph made possible instantaneous communication over long wires. Then a short time later, man discovered how to transmit messages in the form of RADIO WAVES.

As you will learn in this chapter, radio waves are propagated. PROPAGATION means "movement through a medium." This is most easily illustrated by light rays. When a light is turned on in a darkened room, light rays travel from the light bulb throughout the room. When a flashlight is turned on, light rays also radiate from its bulb, but are focused into a narrow beam. You can use these examples to picture how radio waves propagate. Like the light in the room, radio waves may spread out in all directions. They can also be focused (concentrated) like the flashlight, depending upon the need. Radio waves are a form of radiant energy, similar to light and heat. Although they can neither be seen nor felt, their presence can be detected through the use of sensitive measuring devices. The speed at which both forms of waves travel is the same; they both travel at the speed of light.

You may wonder why you can see light but not radio waves, which consist of the same form of energy as light. The reason is that you can only "see" what your eyes can detect. Your eyes can detect radiant energy only within a fixed range of frequencies. Since the frequencies of radio waves are below the frequencies your eyes can detect, you cannot see radio waves.

The theory of wave propagation that we discuss in this module applies to Navy electronic equipment, such as radar, navigation, detection, and communication equipment. We will not discuss these individual systems in this module, but we will explain them in future modules.

Q1. What is propagation?

PRINCIPLES OF WAVE MOTION

All things on the earth—on the land, or in the water—are showered continually with waves of energy. Some of these waves stimulate our senses and can be seen, felt, or heard. For instance, we can see light, hear sound, and feel heat. However, there are some waves that do not stimulate our senses. For

example, radio waves, such as those received by our portable radio or television sets, cannot be seen, heard, or felt. A device must be used to convert radio waves into light (TV pictures) and sound (audio) for us to sense them.

A WAVE can be defined as a DISTURBANCE (sound, light, radio waves) that moves through a MEDIUM (air, water, vacuum). To help you understand what is meant by "a disturbance which moves through a medium," picture the following illustration. You are standing in the middle of a wheat field. As the wind blows across the field toward you, you can see the wheat stalks bending and rising as the force of the wind moves into and across them. The wheat appears to be moving toward you, but it isn't. Instead, the stalks are actually moving back and forth. We can then say that the "medium" in this illustration is the wheat and the "disturbance" is the wind moving the stalks of wheat.

WAVE MOTION can be defined as a recurring disturbance advancing through space with or without the use of a physical medium. Wave motion, therefore, is a means of moving or transferring energy from one point to another point. For example, when sound waves strike a microphone, sound energy is converted into electrical energy. When light waves strike a phototransistor or radio waves strike an antenna, they are likewise converted into electrical energy. Therefore, sound, light, and radio waves are all forms of energy that are moved by wave motion. We will discuss sound waves, light waves, and radio waves later.

Q2. How is a wave defined as it applies to wave propagation?

Q3. What is wave motion?

Q4. What are some examples of wave motion?

WAVE MOTION IN WATER

A type of wave motion familiar to almost everyone is the movement of waves in water. We will explain these waves first to help you understand wave motion and the terms used to describe it.

Basic wave motion can be shown by dropping a stone into a pool of water (see figure 1-1). As the stone enters the water, a surface disturbance is created, resulting in an expanding series of circular waves. Figure 1-2 is a diagram of this action. View A shows the falling stone just an instant before it strikes the water. View B shows the action taking place at the instant the stone strikes the surface, pushing the water that is around it upward and outward. In view C, the stone has sunk deeper into the water, which has closed violently over it causing some spray, while the leading wave has moved outward. An instant later, the stone has sunk out of sight, leaving the water disturbed as shown in view D. Here the leading wave has continued to move outward and is followed by a series of waves gradually diminishing in amplitude. Meanwhile, the disturbance at the original point of contact has gradually subsided.

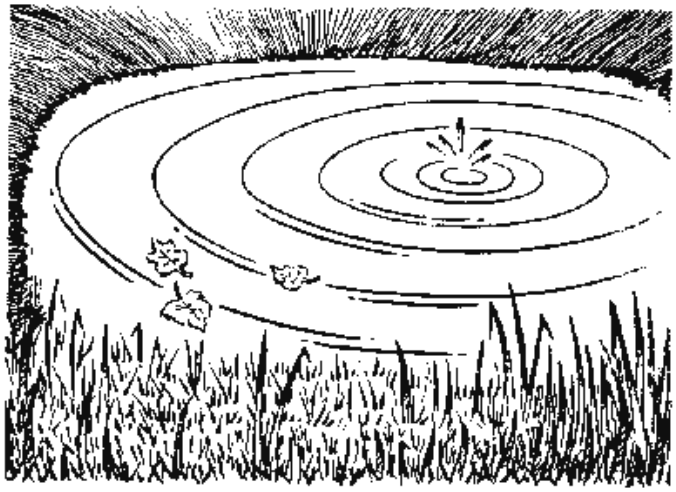


Figure 1-1.—Formation of waves in water.

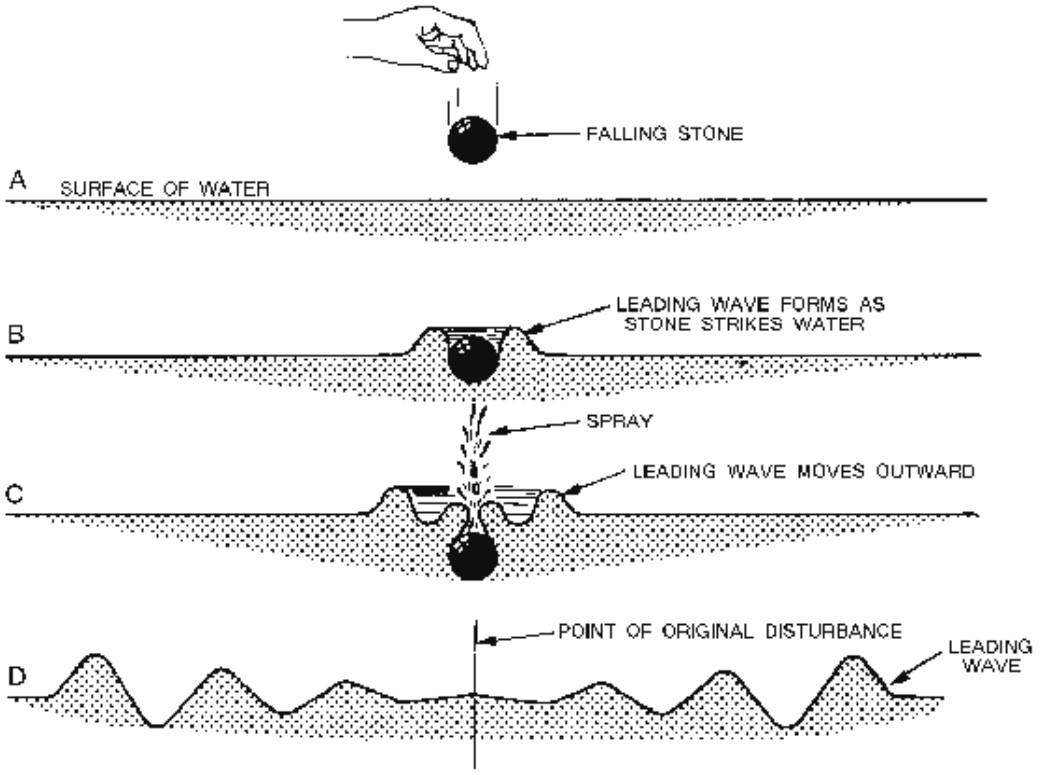


Figure 1-2.—How a falling stone creates wave motion to the surface of water.

In this example, the water is not actually being moved outward by the motion of the waves, but up and down as the waves move outward. The up and down motion is transverse, or at right angles, to the outward motion of the waves. This type of wave motion is called TRANSVERSE WAVE MOTION.

Q5. What type of wave motion is represented by the motion of water?

TRANSVERSE WAVES

To explain transverse waves, we will again use our example of water waves. Figure 1-3 is a cross section diagram of waves viewed from the side. Notice that the waves are a succession of crests and troughs. The wavelength (one 360 degree cycle) is the distance from the crest of one wave to the crest of the next, or between any two similar points on adjacent waves. The amplitude of a transverse wave is half the distance measured vertically from the crest to the trough. Water waves are known as transverse waves because the motion of the water is up and down, or at right angles to the direction in which the waves are traveling. You can see this by observing a cork bobbing up and down on water as the waves pass by; the cork moves very little in a sideways direction. In figure 1-4, the small arrows show the up-and-down direction the cork moves as the transverse wave is set in motion. The direction the wave travels is shown by the large arrow. Radio waves, light waves, and heat waves are examples of transverse waves.

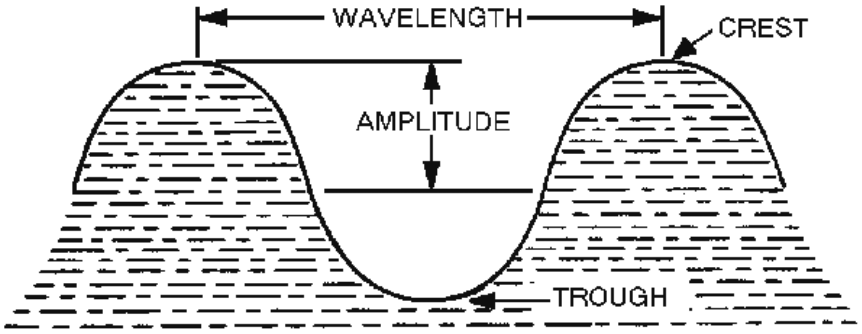


Figure 1-3.—Elements of a wave.

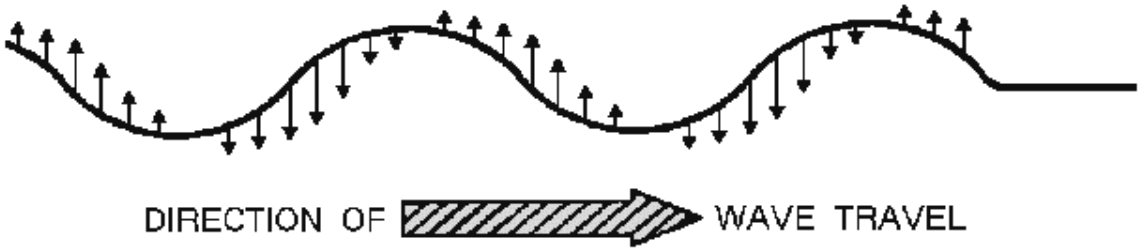


Figure 1-4.—Transverse wave.

LONGITUDINAL WAVES

In the previous discussion, we listed radio waves, light waves, and heat waves as examples of transverse waves, but we did not mention sound waves. Why? Simply because sound waves are LONGITUDINAL WAVES. Unlike transverse waves, which travel at right angles to the direction of propagation, sound waves travel back and forth in the same direction as the wave motion. Therefore, longitudinal waves are waves in which the disturbance takes place in the direction of propagation. Longitudinal waves are sometimes called COMPRESSION WAVES.

Waves that make up sound, such as those set up in the air by a vibrating tuning fork, are longitudinal waves. In figure 1-5, the tuning fork, when struck, sets up vibrations. As the tine moves in an outward direction, the air immediately in front of it is compressed (made more dense) so that its momentary

pressure is raised above that at other points in the surrounding medium (air). Because air is elastic, the disturbance is transmitted in an outward direction as a COMPRESSION WAVE. When the tine returns and moves in the inward direction, the air in front of the tine is rarefied (made less dense or expanded) so that its pressure is lowered below that of the other points in the surrounding air. The rarefied wave is propagated from the tuning fork and follows the compressed wave through the medium (air).

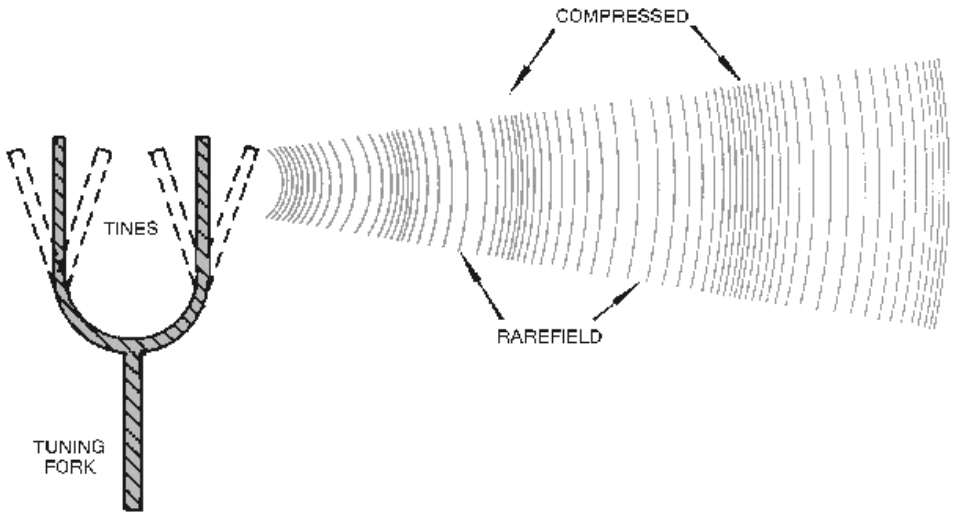


Figure 1-5.—Sound propagation by a tuning fork.

Q6. What are some examples of transverse waves?

Q7. What example of a longitudinal wave was given in the text?

MEDIUM

We have used the term *medium* in describing the motion of waves. Since *medium* is a term that is used frequently in discussing propagation, it needs to be defined so you will understand what a medium is and its application to propagation.

A MEDIUM is the vehicle through which the wave travels from one point to the next. The vehicle that carries a wave can be just about anything. An example of a medium, already mentioned, is air. Air, as defined by the dictionary, is the mixture of invisible, odorless, tasteless gases that surrounds the earth (the atmosphere). Air is made up of molecules of various gases (and impurities). We will call these molecules of air *particles of air* or simply *particles*.

Figure 1-6 will help you to understand how waves travel through air. The object producing the waves is called the SOURCE—a bell in this illustration. The object responding to the waves is called a DETECTOR or RECEIVER—in this case, the human ear. The medium is air, which is the means of conveying the waves from the source to the detector. The source, detector, and medium are all necessary for wave motion and wave propagation (except for electromagnetic waves which require no medium). The waves shown in figure 1-6 are sound waves. As the bell is rung, the particles of air around the bell are compressed and then expanded. This compression and expansion of particles of air set up a wave motion in the air. As the waves are produced, they carry energy from particle to particle through the medium (air) to the detector (ear).

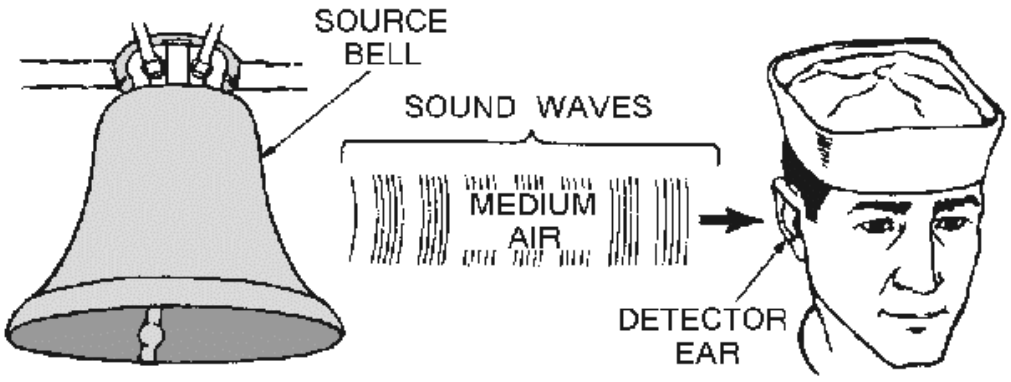


Figure 1-6.—The three elements of sound.

Q8. What are the three requirements for a wave to be propagated?

TERMS USED IN WAVE MOTION

There are a number of special terms concerning waves that you should know. Many of the terms, such as CYCLE, WAVELENGTH, AMPLITUDE, and FREQUENCY were introduced in previous *NEETS* modules. We will now discuss these terms in detail as they pertain to wave propagation. Before we begin our discussion, however, note that in the figure, wave 1 and wave 2 have equal frequency and wavelength but different amplitudes. The REFERENCE LINE (also known as REST POSITION or POINT OF ZERO DISPLACEMENT) is the position that a particle of matter would have if it were not disturbed by wave motion. For example, in the case of the water wave, the reference line is the level of the water when no wave motion is present. With this in mind, let's go on to our discussion of the four terms, as shown in figure 1-7.

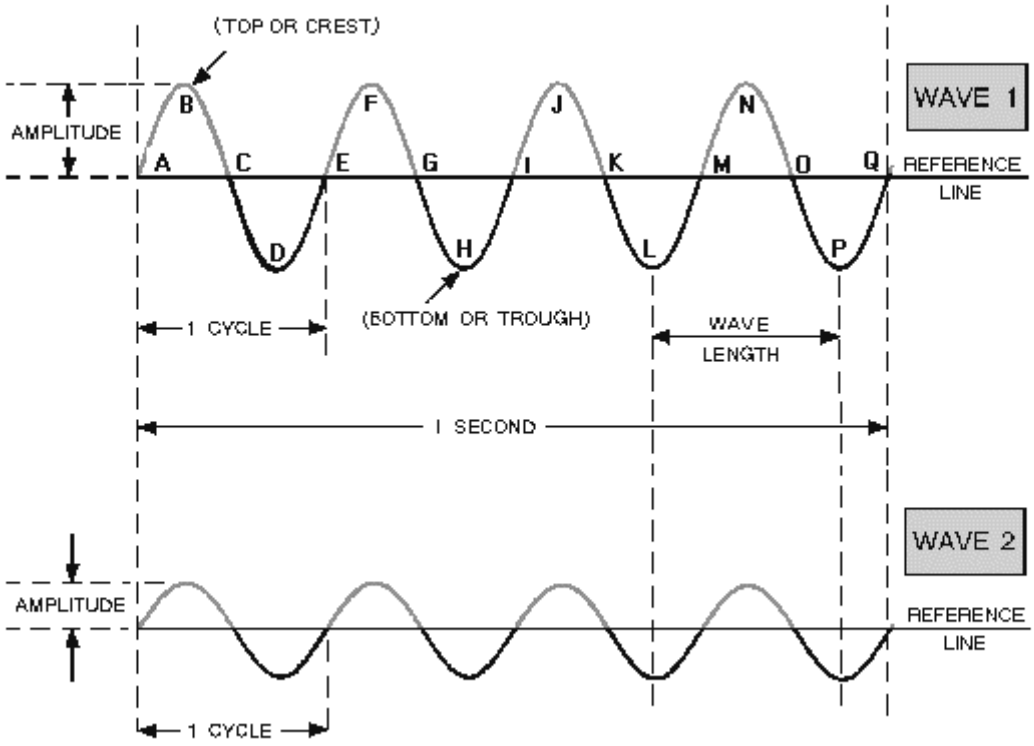


Figure 1-7.—Comparison of waves with different amplitudes.

Cycle

Refer to wave 1 in figure 1-7. Notice how similar it is to the sine wave you have already studied. All transverse waves appear as sine waves when viewed from the side. In figure 1-7, wave 1 has four complete cycles. Points ABCDE comprise one complete cycle having a maximum value above and a maximum value below the reference line. The portion above the reference line (between points A and C) is called a POSITIVE ALTERNATION and the portion below the reference line (between points C and E) is known as a NEGATIVE ALTERNATION. The combination of one complete positive and one complete negative alternation represents one cycle of the wave. At point E, the wave begins to repeat itself with a second cycle completed at point I, a third at point M, etc. The peak of the positive alternation (maximum value above the line) is sometimes referred to as the TOP or CREST, and the peak of the negative alternation (maximum value below the line) is sometimes called the BOTTOM or TROUGH, as depicted in the figure. Therefore, one cycle has one crest and one trough.

Wavelength

A WAVELENGTH is the distance in space occupied by one cycle of a radio wave at any given instant. If the wave could be frozen in place and measured, the wavelength would be the distance from the leading edge of one cycle to the corresponding point on the next cycle. Wavelengths vary from a few hundredths of an inch at extremely high frequencies to many miles at extremely low frequencies; however, common practice is to express wavelengths in meters. Therefore, in figure 1-7 (wave 1), the distance between A and E, or B and F, etc., is one wavelength. The Greek letter lambda (λ) is used to signify wavelength. Why lambda and not "l" or "L"? This is because "L" is used conventionally as the

symbol for inductance, and "l" is used for dimensional length; therefore, λ; is used to indicate the length of waves.

Amplitude

Two waves may have the same wavelength, but the crest of one may rise higher above the reference line than the crest of the other. Compare wave 1 and wave 2 of figure 1-7 again. The height of a wave crest above the reference line is called the AMPLITUDE of the wave. The amplitude of a wave gives a relative indication of the amount of energy the wave transmits. A continuous series of waves, such as A through Q, having the same amplitude and wavelength, is called a train of waves or WAVE TRAIN.

Frequency and Time

Time is an important factor in wave studies. When a wave train passes through a medium, a certain number of individual waves pass a given point in a specific unit of time. For example, if a cork on a water wave rises and falls once every second, the wave makes one complete up-and-down vibration every second. The number of vibrations, or cycles, of a wave train in a unit of time is called the FREQUENCY of the wave train and is measured in HERTZ. If 5 waves pass a point in one second, the frequency of the wave train is 5 cycles per second. In figure 1-7, the frequency of both wave 1 and wave 2 is four cycles per second (cycles per second is abbreviated as cps).

In 1967, in honor of the German physicist Heinrich Hertz, the term HERTZ was designated for use in lieu of the term "cycle per second" when referring to the frequency of radio waves. It may seem confusing that in one place the term "cycle" is used to designate the positive and negative alternations of a wave, but in another instance the term "hertz" is used to designate what appears to be the same thing. The key is the time factor. The term cycle refers to any sequence of events, such as the positive and negative alternations, comprising one cycle of electrical current. The term hertz refers to the number of occurrences that take place in one second.

Q9. What is a cycle?

Q10. What is wavelength (λ)?

CHARACTERISTICS OF WAVE MOTION

The two types of wave motion, transverse and longitudinal, have many of the same characteristics, such as frequency, amplitude, and wavelength. Another important characteristic that these two types of wave motion share is VELOCITY. Velocity of propagation is the rate at which the disturbance travels through the medium, or the velocity with which the crest of the wave moves along. The velocity of the wave depends both on the type of wave (light, sound, or radio) and type of medium (air, water, or metal). If longitudinal waves are plotted as a graph, they appear as transverse waves. This fact is illustrated in figure 1-8.

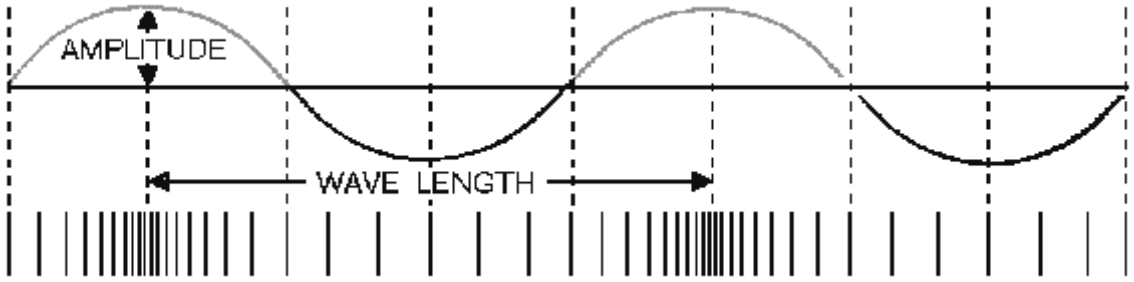


Figure 1-8.—Longitudinal wave represented graphically by a transverse wave.

The frequency of a longitudinal wave, like that of a transverse wave, is the number of complete cycles the wave makes during a specific unit of time. The higher the frequency, the greater is the number of compressions and expansions per unit of time.

In the two types of wave motion described in the preceding discussion, the following quantities are of interest:

a. The PERIOD, which is the time (T) in which one complete vibratory cycle of events occurs,

b. The FREQUENCY OF VIBRATION (f), which is the number of cycles taking place in one second, and

c. The WAVELENGTH, which is the distance the disturbance travels during one period of vibration.

Now, consider the following concept. If a vibrating object makes a certain number of vibrations per second, then 1 second divided by the number of vibrations is equal to the period of time of 1 vibration. In other words, the period, or time, of 1 vibration is the reciprocal of the frequency; thus,

$$\text{time (T) of one vibration} = \frac{1}{\text{frequency (f)}}$$

or

$$T = \frac{1}{f}$$

If you know the velocity of a wave, you can determine the wavelength by dividing the velocity by the frequency. As an equation:

Where:

$$\lambda = \frac{v}{f}$$

λ = wavelength
v = velocity of propagation
f = frequency of vibration

When you use the above equation, be careful to express velocity and wavelength in the proper units of length. For example, in the English system, if the velocity (expressed in feet per second) is divided by the frequency (expressed in cycles per second, or Hz), the wavelength is given in feet per cycle. If the metric system is used and the velocity (expressed in meters per second) is divided by the frequency (expressed in cycles per second), the wavelength is given in meters per cycle. Be sure to express both the wavelength and the frequency in the same units. (Feet per cycle and meters per cycle are normally abbreviated as feet or meters because one wavelength indicates one cycle.) Because this equation holds true for both transverse and longitudinal waves, it is used in the study of both electromagnetic waves and sound waves.

Consider the following example. Two cycles of a wave pass a fixed point every second, and the velocity of the wave train is 4 feet per second. What is the wavelength? The formula for determining wavelength is as follows:

$$\lambda = \frac{v}{f}$$

Where:

λ = wavelength in feet
v = velocity in feet per second
f = frequency in Hz

Given:

v = 4 feet per second
f = 2 Hz

Solution:

$$\lambda = \frac{v}{f}$$

$$\lambda = \frac{4 \text{ feet per second}}{2 \text{ Hz}}$$

$$\lambda = 2 \text{ feet}$$

NOTE: In problems of this kind, be sure NOT to confuse wave velocity with frequency. FREQUENCY is the number of cycles per unit of time (Hz). WAVE VELOCITY is the speed with which a wave train passes a fixed point.

Here is another problem. If a wave has a velocity of 1,100 feet per second and a wavelength of 30 feet, what is the frequency of the wave?

By transposing the general equation:

By transposing the general equation:

$$f = \frac{v}{\lambda}$$

We have the equation:

$$\lambda = \frac{v}{f}$$

Given:

$v = 1{,}100$ feet per second
$\lambda = 30$ feet

Solution:

$$f = \frac{1{,}100 \text{ feet per second}}{30 \text{ feet}}$$

$$f = 36.67\text{Hz}$$

To find the velocity, rewrite the equation as:

$$v = \lambda f$$

Let's work one more problem, this time using the metric system.

Suppose the wavelength is 0.4 meters and the frequency is 12 kHz. What is the velocity?

Use the formula:

velocity = wavelength × frequency ($v = \lambda f$)

Given:

wavelength (λ) = 0.4 meters
frequency (f) = 12kHz

Solution:

$v = \lambda \times f$
$v = 0.4 \text{ meters} \times 12{,}000\text{Hz}$
$v = 4{,}800$ meters per second

Other important characteristics of wave motion are reflection, refraction, diffraction, and the Doppler effect. Big words, but the concept of each is easy to see. For ease of understanding, we will explain the first two characteristics using light waves, and the last two characteristics using sound waves. You should keep in mind that all waves react in a similar manner.

Within mediums, such as air, solids, or gases, a wave travels in a straight line. When the wave leaves the boundary of one medium and enters the boundary of a different medium, the wave changes direction. For our purposes in this module, a boundary is an imaginary line that separates one medium from another.

When a wave passes through one medium and encounters a medium having different characteristics, three things can occur to the wave: (1) Some of the energy can be reflected back into the initial medium; (2) some of the energy can be transmitted into the second medium where it may continue at a different velocity; or (3) some of the energy can be absorbed by the medium. In some cases, all three processes (reflection, transmission, and absorption) may occur to some degree.

Reflection

REFLECTION WAVES are simply waves that are neither transmitted nor absorbed, but are reflected from the surface of the medium they encounter. If a wave is directed against a reflecting surface, such as a mirror, it will reflect or "bounce" from the mirror. Refer to figure 1-9. A wave directed *toward* the surface of the mirror is called the INCIDENT wave. When the wave bounces off of the mirror, it becomes known as the REFLECTED wave. An imaginary line perpendicular to the mirror at the point at which the incident wave strikes the mirror's surface is called the NORMAL, or perpendicular. The angle between the incident wave and the normal is called the ANGLE OF INCIDENCE. The angle between the reflected wave and the normal is called the ANGLE OF REFLECTION.

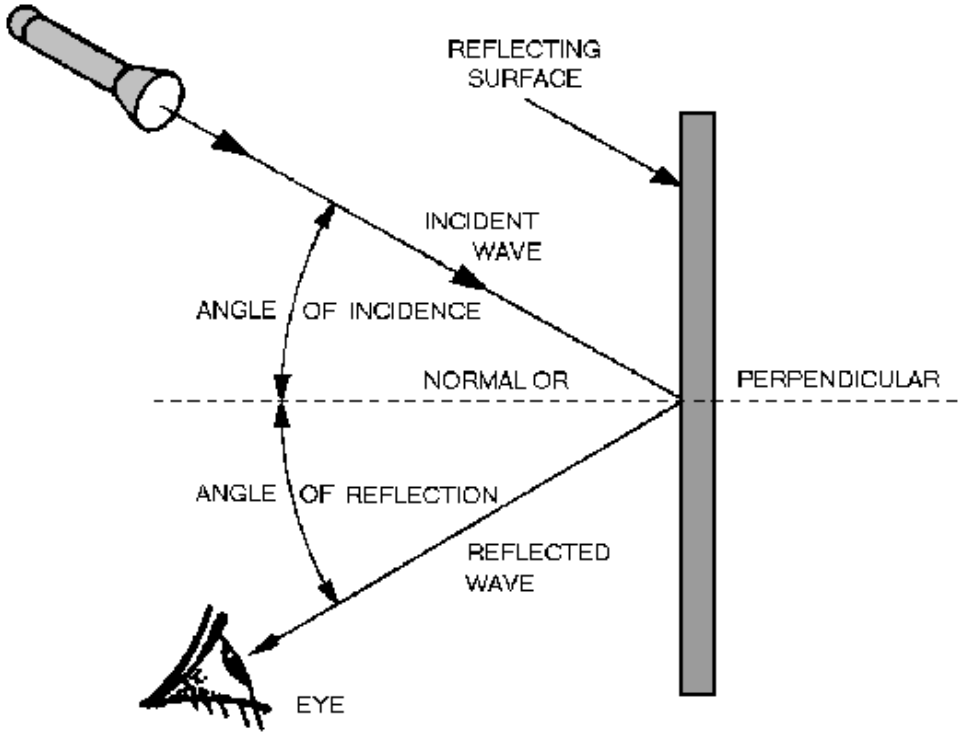


Figure 1-9.—Reflection of a wave.

If the reflecting surface is smooth and polished, the angle between the incident ray and the normal will be the same as the angle between the reflected ray and the normal. This conforms to the law of reflection which states: The angle of incidence is equal to the angle of reflection.

The amount of incident wave energy reflected from a given surface depends on the nature of the surface and the angle at which the wave strikes the surface. As the angle of incidence increases, the amount of wave energy reflected increases. The reflected energy is the greatest when the wave is nearly parallel to the reflecting surface. When the incident wave is perpendicular to the surface, more of the energy is transmitted into the substance and less is reflected. At any incident angle, a mirror reflects almost all of the wave energy, while a dull, black surface reflects very little.

Q11. What is the law of reflection?

Q12. When a wave is reflected from a surface, energy is transferred. When is the transfer of energy greatest?

Q13. When is the transfer of energy minimum?

Refraction

When a wave passes from one medium into another medium that has a different velocity of propagation, a change in the direction of the wave will occur. This changing of direction as the wave enters the second medium is called REFRACTION. As in the discussion of reflection, the wave striking the boundary (surface) is called the INCIDENT WAVE, and the imaginary line perpendicular to the boundary is called the NORMAL. The angle between the incident wave and the normal is called the ANGLE OF INCIDENCE. As the wave passes through the boundary, it is bent either toward or away from the normal. The angle between the normal and the path of the wave through the second medium is the ANGLE OF REFRACTION.

A light wave passing through a block of glass is shown in figure 1-10. The wave moves from point A to B at a constant speed. This is the incident wave. As the wave penetrates the glass boundary at point B, the velocity of the wave is slowed down. This causes the wave to bend toward the normal. The wave then takes the path from point B to C through the glass and becomes BOTH the refracted wave from the top surface and the incident wave to the lower surface. As the wave passes from the glass to the air (the second boundary), it is again refracted, this time away from the normal and takes the path from point C to D. As the wave passes through the last boundary, its velocity increases to the original velocity. As figure 1-10 shows, refracted waves can bend toward or away from the normal. This bending depends on the velocity of the wave through each medium. The broken line between points B and E is the path that the wave would travel if the two mediums (air and glass) had the same density.

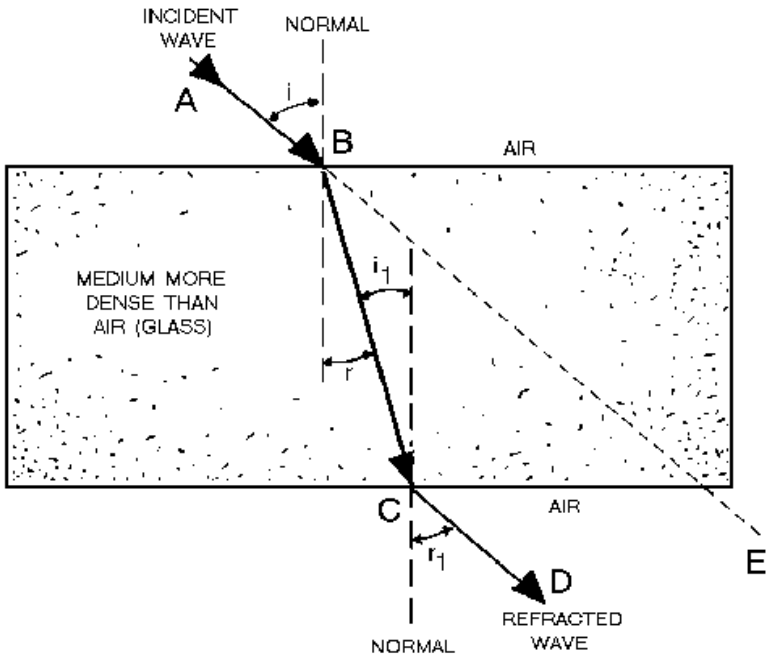


Figure 1-10.—Refraction of a wave.

To summarize what figure 1-10 shows:

1. If the wave passes from a less dense medium to a more dense medium, it is bent toward the normal, and the angle of refraction (r) is less than the angle of incidence (i).

2. If the wave passes from a more dense to a less dense medium, it is bent away from the normal, and the angle of refraction (r_1) is greater than the angle of incidence (i_1).

You can more easily understand refraction by looking at figure 1-11. There is a plowed field in the middle of a parade ground. Think of the incident wave as a company of recruits marching four abreast at an angle across the parade ground to the plowed field, then crossing the plowed field and coming out on the other side onto the parade ground again. As the recruits march diagonally across the parade ground and begin to cross the boundary onto the plowed field, the front line is slowed down. Because the recruits arrive at the boundary at different times, they will begin to slow down at different times (number 1 slows down first and number 4 slows down last in each line). The net effect is a bending action. When the recruits leave the plowed field and reenter the parade ground, the reverse action takes place.

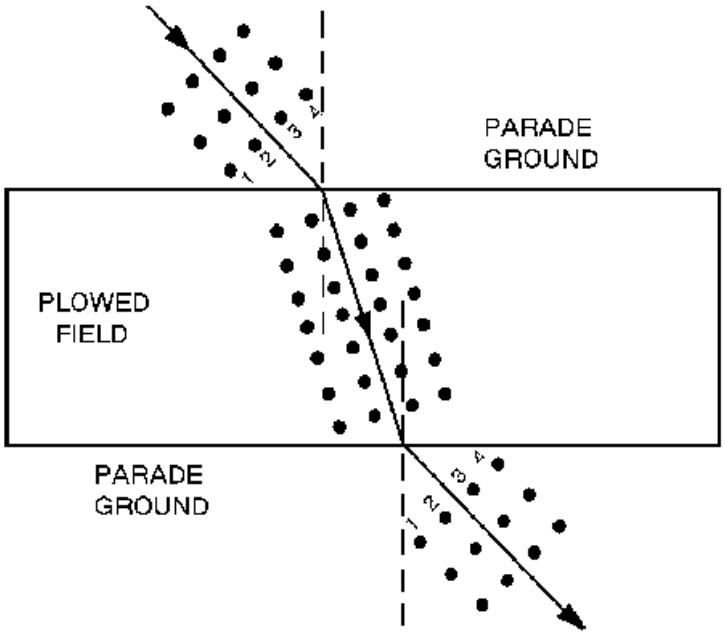


Figure 1-11.—Analogy of refraction.

Q14. A refracted wave occurs when a wave passes from one medium into another medium. What determines the angle of refraction?

Diffraction

DIFFRACTION is the bending of the wave path when the waves meet an obstruction. The amount of diffraction depends on the wavelength of the wave. Higher frequency waves are rarely diffracted in the normal world that surrounds us. Since light waves are high frequency waves, you will rarely see light diffracted. You can, however, observe diffraction in sound waves by listening to music. Suppose you are outdoors listening to a band. If you step behind a solid obstruction, such as a brick wall, you will hear mostly low notes. This is because the higher notes, having short wave lengths, undergo little or no diffraction and pass by or over the wall without wrapping around the wall and reaching your ears. The low notes, having longer wavelengths, wrap around the wall and reach your ears. This leads to the general statement that lower frequency waves tend to diffract more than higher frequency waves. Broadcast band (AM band) radio waves (lower frequency waves) often travel over a mountain to the opposite side from their source because of diffraction, while higher frequency TV and FM signals from the same source tend to be stopped by the mountain.

Doppler Effect

The last, but equally important, characteristic of a wave that we will discuss is the Doppler effect. The DOPPLER EFFECT is the apparent change in frequency or pitch when a sound source moves either toward or away from the listener, or when the listener moves either toward or away from the sound source. This principle, discovered by the Austrian physicist Christian Doppler, applies to all wave motion.

The apparent change in frequency between the source of a wave and the receiver of the wave is because of relative motion between the source and the receiver. To understand the Doppler effect, first assume that the frequency of a sound from a source is held constant. The wavelength of the sound will also remain constant. If both the source and the receiver of the sound remain stationary, the receiver will

hear the same frequency sound produced by the source. This is because the receiver is receiving the same number of waves per second that the source is producing. Now, if either the source or the receiver or both move toward the other, the receiver will perceive a higher frequency sound. This is because the receiver will receive a greater number of sound waves per second and interpret the greater number of waves as a higher frequency sound. Conversely, if the source and the receiver are moving apart, the receiver will receive a smaller number of sound waves per second and will perceive a lower frequency sound. In both cases, the frequency of the sound produced by the source will have remained constant.

For example, the frequency of the whistle on a fast-moving train sounds increasingly higher in pitch as the train is approaching than when the train is departing. Although the whistle is generating sound waves of a constant frequency, and though they travel through the air at the same velocity in all directions, the distance between the approaching train and the listener is decreasing. As a result, each wave has less distance to travel to reach the observer than the wave preceding it. Thus, the waves arrive with decreasing intervals of time between them.

These apparent changes in frequency, called the Doppler effect, affect the operation of equipment used to detect and measure wave energy. In dealing with electromagnetic wave propagation, the Doppler principle is used in equipment such as radar, target detection, weapons control, navigation, and sonar.

Q15. The apparent change in frequency or pitch because of motion is explained by what effect?

SOUND WAVES

The study of sound is important because of the role sound plays in the depth finding equipment (fathometer) and underwater detection equipment (sonar) used by the Navy.

As you know, sound travels through a medium by wave motion. Although sound waves and the electromagnetic waves used in the propagation of radio and radar differ, both types of waves have many of the same characteristics. Studying the principles of sound-wave motion will help you understand the actions of both sound waves and the more complex radio and radar electromagnetic waves. The major differences among sound waves, heat waves, and light waves are (1) their frequencies; (2) their types; the mediums through which they travel; and the velocities at which they travel.

SOUND—WHAT IS IT?

The word SOUND is used in everyday speech to signify a variety of things. One definition of sound is the sensation of hearing. Another definition refers to a stimulus that is capable of producing the sensation of hearing. A third definition limits sound to what is actually heard by the human ear.

In the study of physics, sound is defined as *a range of compression-wave frequencies to which the human ear is sensitive*. For the purpose of this chapter, however, we need to broaden the definition of sound to include compression waves that are not always audible to the human ear. To distinguish frequencies in the audible range from those outside that range, the words SONIC, ULTRASONIC, and INFRASONIC are used. Sounds capable of being heard by the human ear are called SONICS. The normal hearing range extends from about 20 to 20,000 hertz. However, to establish a standard sonic range, the Navy has set an arbitrary upper limit for sonics at 10,000 hertz and a lower limit at 15 hertz. Even though the average person can hear sounds above 10,000 hertz, it is standard practice to refer to sounds above that frequency as *ultrasonic*. Sounds between 15 hertz and 10,000 hertz are called *sonic*, while sounds below 15 hertz are known as *infrasonic* (formerly referred to as subsonic) sounds.

Q16. What term describes sounds capable of being heard by the human ear?

Q17. Are all sounds audible to the human ear? Why?

REQUIREMENTS FOR SOUND

Recall that sound waves are compression waves. The existence of compression waves depends on the transfer of energy. To produce vibrations that become sounds, a mechanical device (the source) must first receive an input of energy. Next, the device must be in contact with a medium that will receive the sound energy and carry it to a receiver. If the device is not in contact with a medium, the energy will not be transferred to a receiver, and there will be no sound.

Thus, three basic elements for transmission and reception of sound must be present before a sound can be produced. They are (1) the source (or transmitter), (2) a medium for carrying the sound (air, water, metal, etc.), and (3) the detector (or receiver).

A simple experiment provides convincing evidence that a medium must be present if sound is to be transferred. In figure 1-12, an electric bell is suspended by rubber bands in a bell jar from which the air can be removed. An external switch is connected from a battery to the bell so the bell may be rung intermittently. As the air is pumped out, the sound from the bell becomes weaker and weaker. If a perfect vacuum could be obtained, and if no sound were conducted out of the jar by the rubber bands, the sound from the bell would be completely inaudible. In other words, sound cannot be transmitted through a vacuum. When the air is admitted again, the sound is as loud as it was at the beginning. This experiment shows that when air is in contact with the vibrating bell, it carries energy to the walls of the jar, which in turn are set in vibration. Thus, the energy passes into the air outside of the jar and then on to the ear of the observer. This experiment illustrates that sound cannot exist in empty space (or a vacuum).

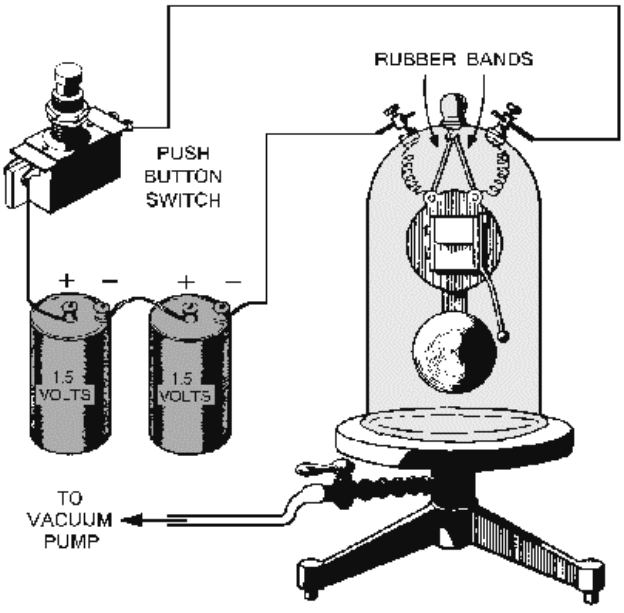


Figure 1-12.—No air, no sound.

Any object that moves rapidly back and forth, or vibrates, and thus disturbs the medium around it may be considered a source for sound. Bells, speakers, and stringed instruments are familiar sound sources.

The material through which sound waves travel is called the *medium*. The density of the medium determines the ease, distance, and speed of sound transmission. The higher the density of the medium, the slower sound travels through it.

The detector acts as the receiver of the sound wave. Because it does not surround the source of the sound wave, the detector absorbs only part of the energy from the wave and sometimes requires an amplifier to boost the weak signal.

As an illustration of what happens if one of these three elements is not present, let's refer to our experiment in which a bell was placed in a jar containing a vacuum. You could see the bell being struck, but you could hear no sound because there was no medium to transmit sound from the bell to you. Now let's look at another example in which the third element, the detector, is missing. You see a source (such as an explosion) apparently producing a sound, and you know the medium (air) is present, but you are too far away to hear the noise. Thus, as far as you are concerned, there is no detector and, therefore, no sound. We must assume, then, that sound can exist only when a source transmits sound through a medium, which passes it to a detector. Therefore, in the absence of any one of the basic elements (source, medium, detector) there can be NO sound.

Q18. Sound waves transmitted from a source are sometimes weak when they reach the detector. What instrument is needed to boost the weak signal?

TERMS USED IN SOUND WAVES

Sound waves vary in length according to their frequency. A sound having a long wavelength is heard at a low pitch (low frequency); one with a short wavelength is heard at a high pitch (high frequency). A complete wavelength is called a cycle. The distance from one point on a wave to the corresponding point on the next wave is a wavelength. The number of cycles per second (hertz) is the frequency of the sound. The frequency of a sound wave is also the number of vibrations per second produced by the sound source.

Q19. What are the three basic requirements for sound?

CHARACTERISTICS OF SOUND

Sound waves travel at great distances in a very short time, but as the distance increases the waves tend to spread out. As the sound waves spread out, their energy simultaneously spreads through an increasingly larger area. Thus, the wave energy becomes weaker as the distance from the source is increased.

Sounds may be broadly classified into two general groups. One group is NOISE, which includes sounds such as the pounding of a hammer or the slamming of a door. The other group is musical sounds, or TONES. The distinction between noise and tone is based on the regularity of the vibrations, the degree of damping, and the ability of the ear to recognize components having a musical sequence. You can best understand the physical difference between these kinds of sound by comparing the waveshape of a musical note, depicted in view A of figure 1-13, with the waveshape of noise, shown in view B. You can see by the comparison of the two waveshapes, that noise makes a very irregular and haphazard curve and a musical note makes a uniform and regular curve.

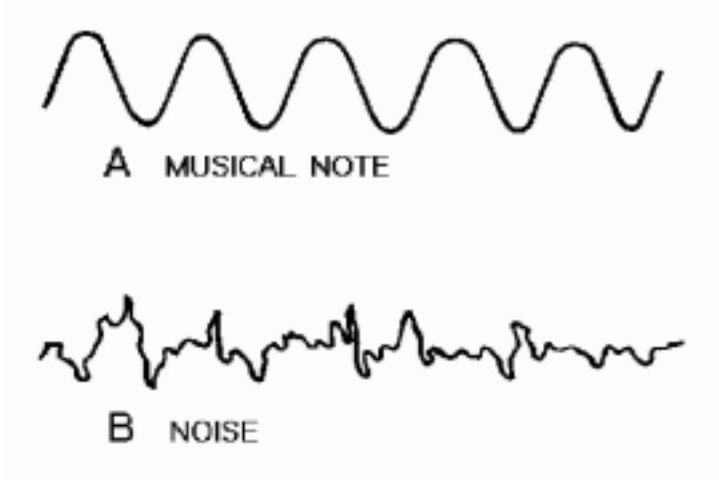


Figure 1-13.—Musical sound versus noise.

Sound has three basic characteristics: pitch, intensity, and quality. Each of these three characteristics is associated with one of the properties of the source or the type of waves which it produces. The pitch depends upon the frequency of the waves; the intensity depends upon the amplitude of the waves; and the quality depends upon the form of the waves. With the proper combination of these characteristics, the tone is pleasant to the ear. With the wrong combination, the sound quality turns into noise.

The Pitch of Sound

The term PITCH is used to describe the frequency of a sound. An object that vibrates many times per second produces a sound with a high pitch, as with a police whistle. The slow vibrations of the heavier strings of a violin cause a low-pitched sound. Thus, the frequency of the wave determines pitch. When the frequency is low, sound waves are long; when it is high, the waves are short. A sound can be so high in frequency that the waves reaching the ear cannot be heard. Likewise, some frequencies are so low that the eardrums do not convert them into sound. The range of sound that the human ear can detect varies with each individual.

The Intensity of Sound

The intensity of sound, at a given distance, depends upon the amplitude of the waves. Thus, a tuning fork gives out more energy in the form of sound when struck hard than when struck gently. You should remember that when a tuning fork is struck, the sound is omnidirectional (heard in all directions), because the sound waves spread out in all directions, as shown in figure 1-14. You can see from the figure that as the distance between the waves and the sound source increases, the energy in each wave spreads over a greater area; hence, the intensity of the sound decreases. The speaking tubes sometimes used aboard a ship prevent the sound waves from spreading in all directions by concentrating them in one desired direction (unidirectional), producing greater intensity. Therefore, the sound is heard almost at its original intensity at the opposite end of the speaking tube. The unidirectional megaphone and the directional loudspeaker also prevent sound waves from spreading in all directions.

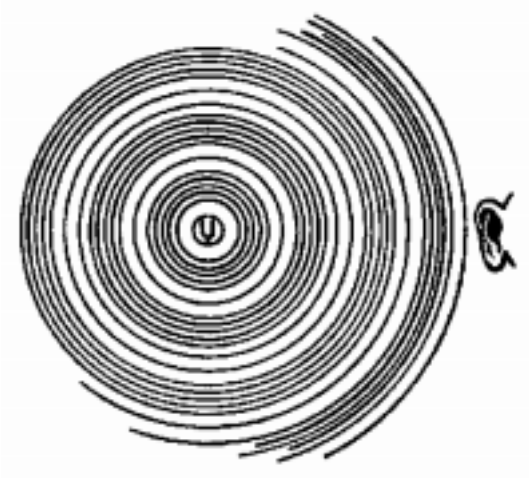


Figure 1-14.—Sound waves spread in all directions.

Sound intensity and loudness are often mistakenly interpreted as having the same meaning. Although they are related, they are not the same. Sound INTENSITY is a measure of the sound energy of a wave. LOUDNESS, on the other hand, is the sensation the intensity (and sometimes frequency) the sound wave produces on the ear. Increasing the intensity causes an increase in loudness but not in a direct proportion. For instance, doubling the loudness of a sound requires about a tenfold increase in the intensity of the sound.

Sound Quality

Most sounds, including musical notes, are not pure tones. They are a mixture of different frequencies (tones). A tuning fork, when struck, produces a pure tone of a specific frequency. This pure tone is produced by regular vibrations of the source (tines of the tuning fork). On the other hand, scraping your fingernails across a blackboard only creates noise, because the vibrations are irregular. Each individual pipe of a pipe organ is similar to a tuning fork, and each pipe produces a tone of a specific frequency. But sounding two or more pipes at the same time produces a complex waveform. A tone that closely imitates any of the vowel sounds can be produced by selecting the proper pipes and sounding them at the same time. Figure 1-15 illustrates the combining of two pure tones to make a COMPLEX WAVE.

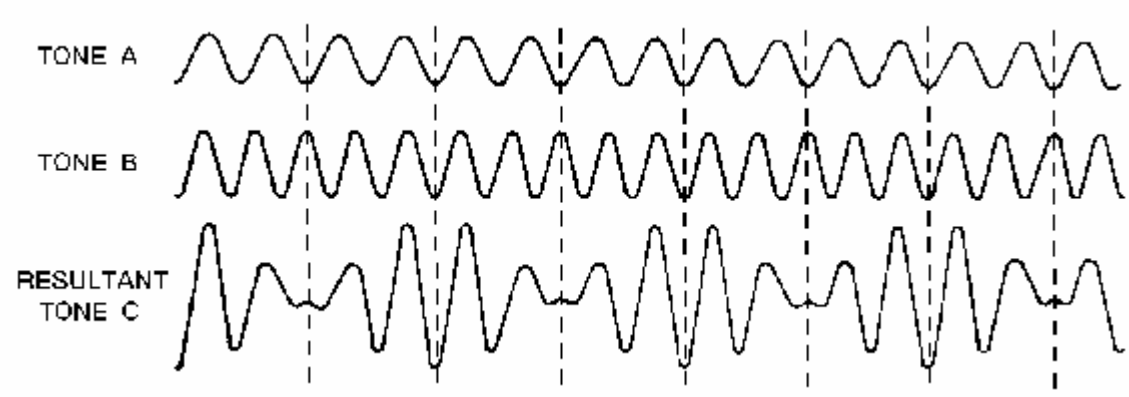


Figure 1-15.—Combination of tones.

The QUALITY of a sound depends on the complexity of its sound waves, such as the waves shown in tone C of figure 1-15. Almost all sounds (musical and vocal included) have complicated (complex)

waveforms. Tone A is a simple wave of a specific frequency that can be produced by a tuning fork, piano, organ, or other musical instrument. Tone B is also a simple wave but at a different frequency. When the two tones are sounded together, the complex waveform in tone C is produced. Note that tone C has the same frequency as tone A with an increase in amplitude. The human ear could easily distinguish between tone A and tone C because of the quality. Therefore, we can say that quality distinguishes tones of like pitch and loudness when sounded on different types of musical instruments. It also distinguishes the voices of different persons.

Q20. What are the two general groups of sound?

Q21. What are the three basic characteristics of sound?

Q22. What is the normal audible range of the human ear?

Q23. What is intensity as it pertains to sound?

Q24. What characteristic of sound enables a person to distinguish one musical instrument from another, if they are all playing the same note?

ELASTICITY AND DENSITY AND VELOCITY OF TRANSMISSION

Sound waves travel through any medium to a velocity that is controlled by the medium. Varying the frequency and intensity of the sound waves will not affect the speed of propagation. The ELASTICITY and DENSITY of a medium are the two basic physical properties that govern the velocity of sound through the medium.

Elasticity is the ability of a strained body to recover its shape after deformation, as from a vibration or compression. The measure of elasticity of a body is the force it exerts to return to its original shape.

The *density* of a medium or substance is the mass per unit volume of the medium or substance. Raising the temperature of the medium (which decreases its density) has the effect of increasing the velocity of sound through the medium.

The velocity of sound in an elastic medium is expressed by the formula:

$$v = \sqrt{\frac{E}{d}}$$

Even though solids such as steel and glass are far more dense than air, their elasticity's are so much greater that the velocities of sound in them are 15 times greater than the velocity of sound in air. Using elasticity as a rough indication of the speed of sound in a given medium, we can state as a general rule that *sound travels faster in harder materials* (such as steel), *slower in liquids, and slowest in gases*. Density has the opposite effect on the velocity of sound, that is, with other factors constant, a denser material (such as lead) passes sound slower.

At a given temperature and atmospheric pressure, all sound waves travel in air at the same speed. Thus the velocity that sound will travel through air at 32º F (0º C) is 1,087 feet per second. But for practical purposes, the speed of sound in air may be considered as 1,100 feet per second. Table 1-1 gives a comparison of the velocity of sound in various mediums.

Table 1-1.—Comparison of Velocity of Sound in Various Mediums

MEDIUM	TEMPERATURE		VELOCITY
	ºF	ºC	(FT/SEC)
AIR	32	0	1,087
AIR	68	20	1,127
ALUMINUM	68	20	16,700
CARBON DIOXIDE	32	0	856
FRESH WATER	32	0	4,629
FRESH WATER	68	20	4,805
HYDROGEN	32	0	4,219
LEAD	32	20	4,030
SALT WATER	32	0	4,800
SALT WATER	68	20	4,953
STEEL	32	0	16,410
STEEL	68	20	16,850

Q25. How does density and temperature affect the velocity of sound?

ACOUSTICS

The science of sound is called ACOUSTICS. This subject could fill volumes of technical books, but we will only scratch the surface in this chapter. We will present important points that you will need for a better understanding of sound waves.

Acoustics, like sound, relates to the sense of hearing. It also deals with the production, control, transmission, reception, and the effects of sound. For the present, we are concerned only with the last relationship—the effects of sound. These same effects will be used throughout your study of wave propagation.

Echo

An ECHO is the reflection of the original sound wave as it bounces off a distant surface. Just as a rubber ball bounces back when it is thrown against a hard surface, sound waves also bounce off most surfaces. As you have learned from the study of the law of conservation of energy, a rubber ball never bounces back with as much energy as the initial bounce. Similarly, a reflected sound wave is not as loud as the original sound wave. In both cases, some of the energy is absorbed by the reflecting surface. Only a portion of the original sound is reflected, and only a portion of the reflected sound returns to the listener. For this reason, an echo is never as loud as the original sound.

Sound reflections (echoes) have many applications in the Navy. The most important of these applications can be found in the use of depth finding equipment (the fathometer) and sonar. The fathometer sends sound-wave pulses from the bottom of the ship and receives echoes from the ocean floor to indicate the depth of the ocean beneath the ship. The sonar transmits a pulse of sound energy and receives the echo to indicate range and bearing of objects or targets in the ocean depths.

Refraction

When sound waves traveling at different velocities pass obliquely (at an angle) from one medium into another, the waves are refracted; that is, their line of travel is bent. Refraction occurs gradually when one part of a sound wave is traveling faster than the other parts. For example, the wind a few feet above

the surface of the earth has a greater velocity than that near the surface because friction retards the lower layers (see figure 1-16). The velocity of the wind is added to the velocity of the sound through the air. The result is that the upper portion of the sound wave moves faster than the lower portion and causes a gradual change in the direction of travel of the wave. Refraction causes sound to travel farther with the wind than against it.

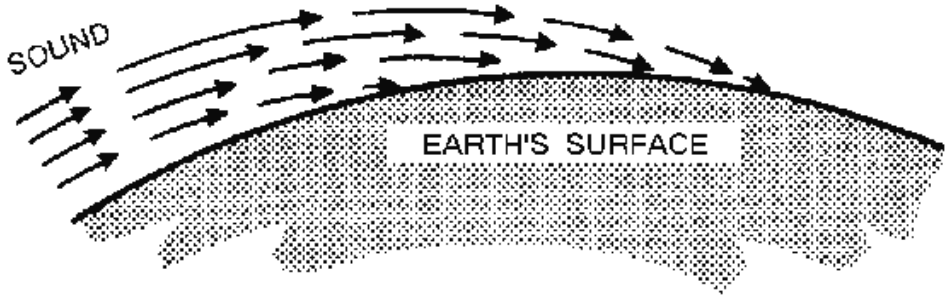


Figure 1-16.—Refraction of sound.

Reverberation

In empty rooms or other confined spaces, sound may be reflected several times to cause what is known as reverberation. REVERBERATION is the multiple reflections of sound waves. Reverberations seem to prolong the time during which a sound is heard. Examples of this often occur in nature. For instance, the discharge of lightning causes a sharp, quick sound. By the time this sound has reached the ears of a distant observer, it is usually drawn out into a prolonged roar by reverberations that we call thunder. A similar case often arises with underwater sound equipment. Reverberations from nearby points may continue for such a long time that they interfere with the returning echoes from targets.

Interference

Any disturbance, man-made or natural, that causes an undesirable response or the degradation of a wave is referred to as INTERFERENCE.

Two sound waves moving simultaneously through the same medium will advance independently, each producing a disturbance as if the other were not present. If the two waves have the same frequency—in phase with each other—and are moving in the same direction, they are additive and are said to interfere constructively. If the two waves have the same frequency and are moving in the same direction, but out of phase with each other, they are subtractive and are said to interfere destructively. If these two subtractive waves have equal amplitudes, the waves cancel each other. This addition or subtraction of waves is often called interference.

Resonance

At some time during your life you probably observed someone putting his or her head into an empty barrel or other cavity and making noises varying in pitch. When that person's voice reached a certain pitch, the tone produced seemed much louder than the others. The reason for this phenomenon is that at that a certain pitch the frequency of vibrations of the voice matched the resonant (or natural) frequency of the cavity. The resonant frequency of a cavity is the frequency at which the cavity body will begin to vibrate and create sound waves. When the resonant frequency of the cavity was reached, the sound of the voice was reinforced by the sound waves created by the cavity, resulting in a louder tone.

This phenomenon occurs whenever the frequency of vibrations is the same as the natural frequency of a cavity, and is called RESONANCE.

Noise

The most complex sound wave that can be produced is noise. Noise has no tonal quality. It distracts and distorts the sound quality that was intended to be heard. NOISE is generally an unwanted disturbance caused by spurious waves originating from man-made or natural sources, such as a jet breaking the sound barrier, or thunder.

Q26. What term is used in describing the science of sound?

Q27. A sound wave that is reflected back toward the source is known as what type of sound?

Q28. What is the term for multiple reflections of sound waves?

Q29. A cavity that vibrates at its natural frequency produces a louder sound than at other frequencies. What term is used to describe this phenomenon?

Q30. What do we call a disturbance that distracts or distorts the quality of sound?

LIGHT WAVES

Technicians maintain equipment that use frequencies from one end of the electromagnetic spectrum to the other—from low-frequency radio waves to high-frequency X-rays and cosmic rays. Visible light is a small but very important part of this electromagnetic spectrum.

Most of the important terms that pertain to the behavior of waves, such as reflection, refraction, diffraction, etc., were discussed earlier in this chapter. We will now discuss how these terms are used in understanding light and light waves. The relationship between light and light waves (rays) is the same as sound and sound waves.

Light is a form of energy. It can be produced by various means (mechanical, electrical, chemical, etc.). We can see objects because the light rays they give off or reflect reach our eyes. If the object is the source of light energy, it is called luminous. If the object is not the source of light but reflects light, it is called an illuminated body.

PROPAGATION OF LIGHT

The exact nature of light is not fully understood, although scientists have been studying the subject for many centuries. Some experiments seem to show that light is composed of tiny particles, and some suggest that it is made up of waves.

One theory after another attracted the approval and acceptance of physicists. Today, some scientific phenomena can be explained only by the wave theory and others only by the particle theory. Physicists, constantly searching for some new discovery that would bring these two theories into agreement, gradually have come to accept a theory that combines the principles of the two theories.

According to the view now generally accepted, light is a form of electromagnetic radiation; that is, light and similar forms of radiation are made up of moving electric and magnetic fields. These two fields will be explained thoroughly later in this chapter.

ELECTROMAGNETIC THEORY OF LIGHT

James Clark Maxwell, a brilliant Scottish scientist Of the middle l9th century, showed, by constructing an oscillating electrical circuit, that electromagnetic waves could move through empty space. Light eventually was proved to be electromagnetic.

Current light theory says that light is made up of very small packets of electromagnetic energy called PHOTONS (the smallest unit of radiant energy). These photons move at a constant speed in the medium through which they travel. Photons move at a faster speed through a vacuum than they do in the atmosphere, and at a slower speed through water than air.

The electromagnetic energy of light is a form of electromagnetic radiation. Light and similar forms of radiation are made up of moving electric and magnetic forces and move as waves. Electromagnetic waves move in a manner similar to the waves produced by the pebble dropped in the pool of water discussed earlier in this chapter. The transverse waves of light from a light source spread out in expanding circles much like the waves in the pool. However, the waves in the pool are very slow and clumsy in comparison with light, which travels approximately 186,000 miles per second.

Light radiates from its source in all directions until absorbed or diverted by some substance (fig. 1-17). The lines drawn from the light source (a light bulb in this instance) to any point on one of these waves indicate the direction in which the waves are moving. These lines, called radii of the spheres, are formed by the waves and are called LIGHT RAYS.

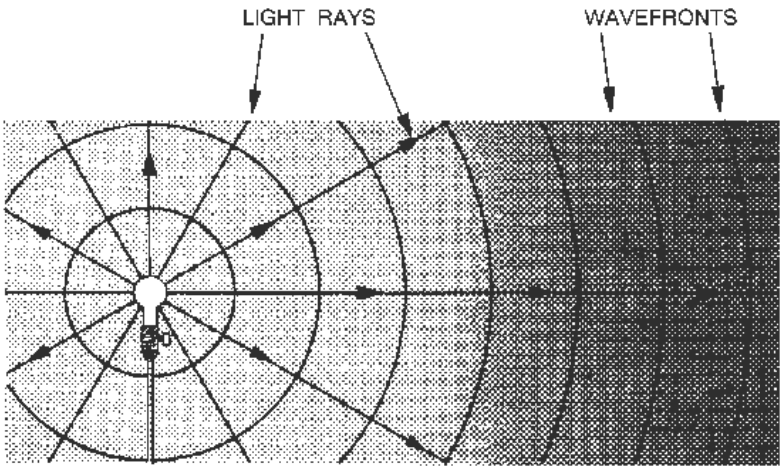


Figure 1-17.—Waves and radii from a nearby light source.

Although single rays of light do not exist, light "rays" as used in illustrations are a convenient method used to show the direction in which light is traveling at any point.

A large volume of light is called a beam; a narrow beam is called a pencil; and the smallest portion of a pencil is called a light ray. A ray of light, can be illustrated as a straight line. This straight line drawn from a light source will represent an infinite number of rays radiating in all directions from the source.

Q31. What are three means of producing light?

Q32. What is the smallest unit of radiant energy?

FREQUENCIES AND WAVELENGTHS

Compared to sound waves, the frequency of light waves is very high and the wavelength is very short. To measure these wavelengths more conveniently, a special unit of measure called an ANGSTROM UNIT, or more usually, an ANGSTROM (Å) was devised. Another common unit used to measure these waves is the millimicron (mμ), which is one millionth of a millimeter. One mF equals ten angstroms. One angstrom equals 1055^{-10}m.

Q33. What unit is used to measure the different wavelengths of light?

FREQUENCIES AND COLOR

For our discussion of light wave waves, we will use the millimicron measurement. The wavelength of a light determines the color of the light. Figure 1-18 indicates that light with a wavelength of 700 millimicrons is red, and that light with a wavelength of 500 millimicrons is blue-green. This illustration shows approximate wavelengths of the different colors in the visible spectrum. In actual fact, the color of light depends on its frequency, not its wavelength. However, light is measured in wavelengths.

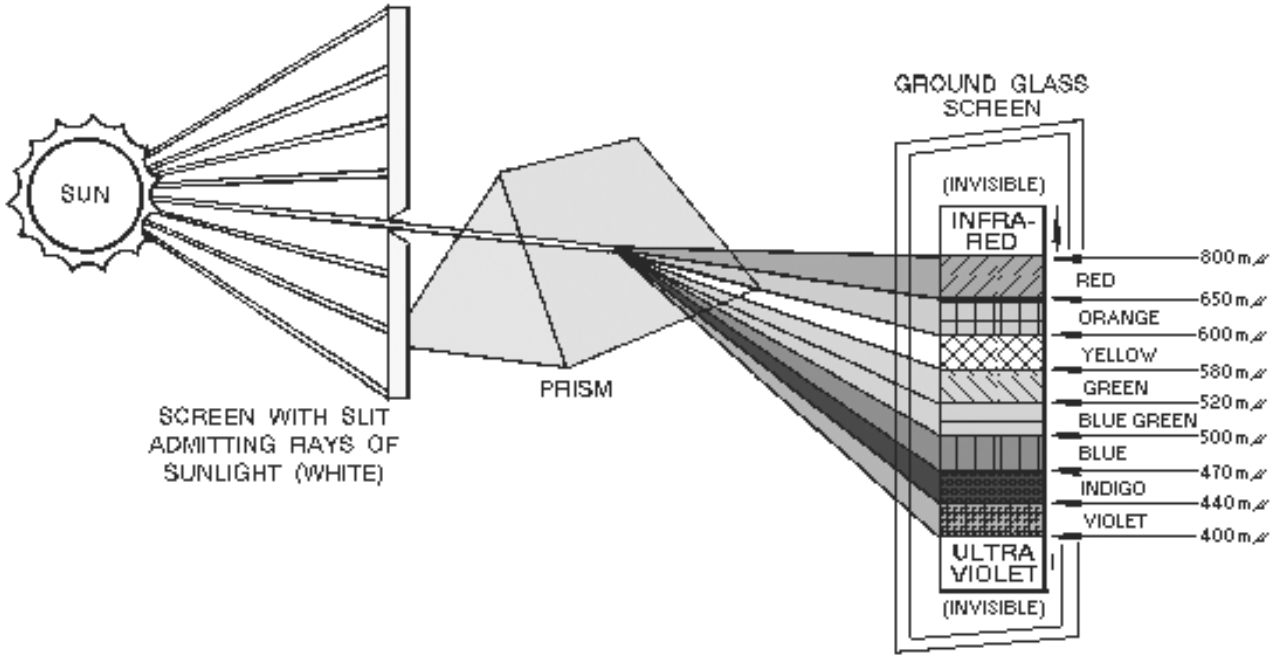


Figure 1-18.—Use of a prism to split white light into different colors.

When the wavelength of 700 millimicrons is measured in a medium such as air, it produces the color red, but the same wave measured in a different medium will have a different wavelength. When red light which has been traveling in air enters glass, it loses speed. Its wavelength becomes shorter or compressed, but it continues to be red. This illustrates that the color of light depends on frequency and not on wavelength. The color scale in figure 1-18 is based on the wavelengths in air.

When a beam of white light (sunlight) is passed through a PRISM, as shown in figure 1-18, it is refracted and dispersed (the phenomenon is known as DISPERSION) into its component wavelengths. Each of these wavelengths causes a different reaction of the eye, which sees the various colors that compose the visible spectrum. The visible spectrum is recorded as a mixture of red, orange, yellow, green, blue, indigo, and violet. White light results when the PRIMARIES (red, green, and blue) are mixed

together in overlapping beams of light. (NOTE: These are not the primaries used in mixing pigments, such as in paint.) Furthermore, the COMPLEMENTARY or SECONDARY colors (magenta, yellow, and cyan) may be shown with equal ease by mixing any two of the primary colors in overlapping beams of light. Thus, red and green light mixed in equal intensities will make yellow light; green and blue will produce cyan (blue-green light); and blue and red correctly mixed will produce magenta (a purplish red light).

LIGHT AND COLOR

All objects absorb some of the light that falls on them. An object appears to be a certain color because it absorbs all of the light waves except those whose frequency corresponds to that particular color. Those waves are reflected from the surface, strike your eye, and cause you to see the particular color. The color of an object therefore depends on the frequency of the electromagnetic wave reflected.

LUMINOUS BODIES

Certain bodies, such as the sun, a gas flame, and an electric light filament, are visible because they are light sources. They are called SELF-LUMINOUS bodies. Objects other than self-luminous bodies become visible only when they are in the presence of light from luminous bodies.

Most NONLUMINOUS bodies are visible because they diffuse or reflect the light that falls on them. A good example of a nonluminous diffusing body is the moon, which shines only because the sunlight falling onto its surface is diffused.

Black objects do not diffuse or reflect light. They are visible only when outlined against a background of light from a luminous or diffusing body.

PROPERTIES OF LIGHT

When light waves, which travel in straight lines, encounter any substance, they are either transmitted, refracted, reflected, or absorbed. This is illustrated in figure 1-19. When light strikes a substance, some absorption and some reflection always take place. No substance completely transmits, reflects, or absorbs all of the light rays that reach its surface. Substances that transmit almost all the light waves that fall upon them are said to be TRANSPARENT. A transparent substance is one through which you can see clearly. Clear glass is transparent because it transmits light rays without diffusing them (view A of figure 1-20). There is no known perfectly transparent substance, but many substances are nearly so. Substances through which some light rays can pass but through which objects cannot be seen clearly because the rays are diffused are called TRANSLUCENT (view B of figure 1-20). The frosted glass of a light bulb and a piece of oiled paper are examples of translucent materials. Substances that do not transmit any light rays are called OPAQUE (view C of figure 1-20). Opaque substances can either reflect or absorb all of the light rays that fall upon them.

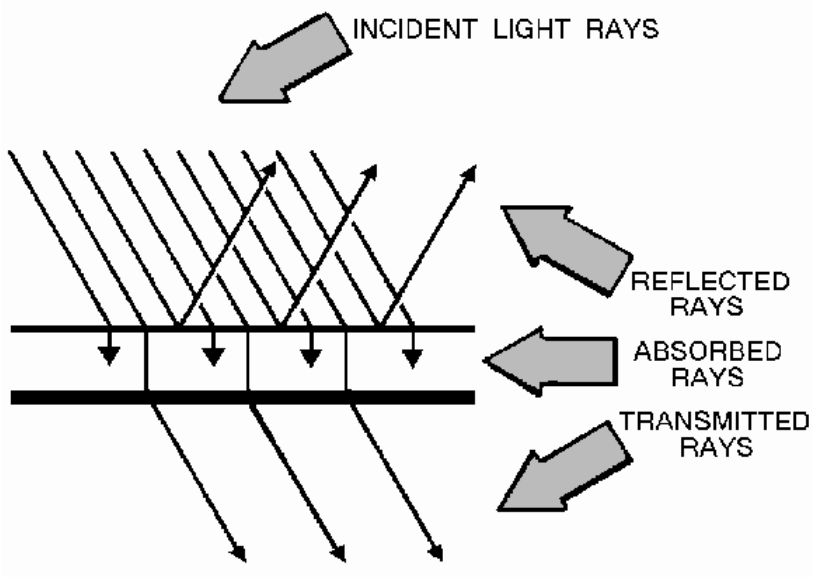


Figure 1-19.—Light waves reflected, absorbed, and transmitted.

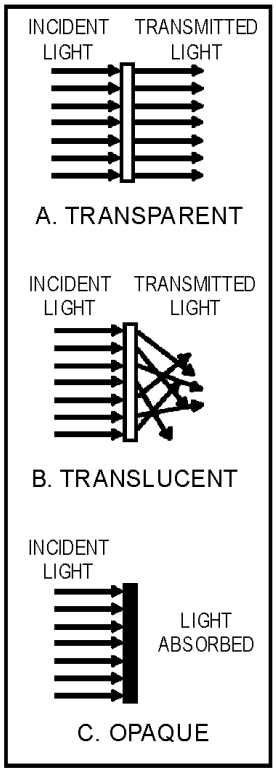


Figure 1-20.—Transparent, translucent, and opaque substances.

Q34. What are the three primary colors of light?

Q35. What are the three secondary colors of light?

Q36. White light falls upon a dull, rough, dark-brown object. Will the light primarily be reflected, diffused, or absorbed by the object?

Q37. What color will be emitted by a dull, rough, black object when white light falls upon it?

Q38. A substance that transmits light but through which an object cannot be seen clearly is known as what kind of substance?

Speed of Light

You probably have heard people say, "quick as lightning" or "fast as light" to describe rapid motion; nevertheless, it is difficult to realize how fast light actually travels. Not until recent years have scientists been able to measure accurately the speed of light.

Prior to the middle 17th century, scientists thought that light required no time at all to pass from the source to the observer. Then in 1675, Ole Roemer, a Danish astronomer, discovered that light travels approximately 186,000 miles per second in space. At this velocity, a light beam can circle the earth 7 1/2 times in one second.

The speed of light depends on the medium through which the light travels. In empty space, the speed is 186,000 (1.86×10^5) miles per second. It is almost the same in air. In water, it slows down to approximately 140,000 (1.4×10^5) miles per second. In glass, the speed of light is 124,000 (1.24×10^5) miles per second. In other words, the speed of light decreases as the density of the substance through which the light passes increases.

The velocity of light, which is the same as the velocity of other electromagnetic waves, is considered to be constant, at 186,000 miles per second. If expressed in meters, it is 300,000,000 meters per second.

Reflection of Light

Light waves obey the law of reflection in the same manner as other types of waves. Consider the straight path of a light ray admitted through a narrow slit into a darkened room. The straight path of the beam is made visible by illuminated dust particles suspended in the air. If the light beam is made to fall onto the surface of a mirror or other reflecting surface, however, the direction of the beam changes sharply. The light can be reflected in almost any direction depending on the angle at which the mirror is held.

As shown earlier in figure 1-9, if a light beam strikes a mirror, the angle at which the beam is reflected depends on the angle at which it strikes the mirror. The beam approaching the mirror is the INCIDENT or striking beam, and the beam leaving the mirror is the REFLECTED beam.

The term "reflected light" simply refers to light waves that are neither transmitted nor absorbed, but are thrown back from the surface of the medium they encounter.

You will see this application used in our discussion of radio waves (chapter 2) and antennas (chapter 4).

Q39. At what speed does light travel?

Refraction of Light

The change of direction that occurs when a ray of light passes from one transparent substance into another of different density is called refraction. Refraction is due to the fact that light travels at various

speeds in different transparent substances. For example, water never appears as deep as it really is, and objects under water appear to be closer to the surface than they really are. A bending of the light rays causes these impressions.

Another example of refraction is the apparent bending of a spoon when it is immersed in a cup of water. The bending seems to take place at the surface of the water, or exactly at the point where there is a change of density. Obviously, the spoon does not bend from the pressure of the water. The light forming the image of the spoon is bent as it passes from the water (a medium of high density) to the air (a medium of comparatively low density).

Without refraction, light waves would pass in straight lines through transparent substances without any change of direction. Refer back to figure 1-10, which shows refraction of a wave. As you can see, all rays striking the glass at any angle other than perpendicular are refracted. However, the perpendicular ray, which enters the glass normal to the surface, continues through the glass and into the air in a straight line no refraction takes place.

Diffusion of Light

When light is reflected from a mirror, the angle of reflection of each ray equals the angle of incidence. When light is reflected from a piece of plain white paper, however, the reflected beam is scattered, or DIFFUSED, as shown in figure 1-21. Because the surface of the paper is not smooth, the reflected light is broken up into many light beams that are reflected in all directions.

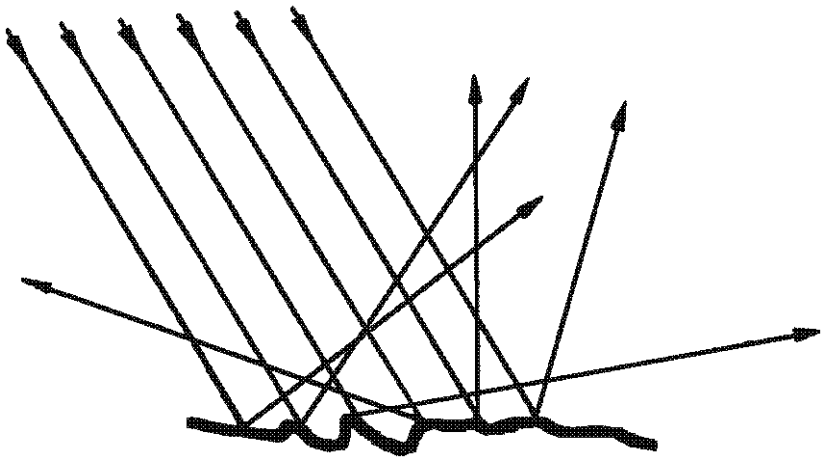


Figure 1-21.—Diffusion of light.

Absorption of Light

You have just seen that a light beam is reflected and diffused when it falls onto a piece of white paper. If a light beam falls onto a piece of black paper, the black paper absorbs most of the light rays and very little light is reflected from the paper. If the surface on which the light beam falls is perfectly black, there is no reflection; that is, the light is totally absorbed. No matter what kind of surface light falls on, however, some of the light is absorbed.

Q40. A light wave enters a sheet of glass at a perfect right angle to the surface. Is the majority of the wave reflected, refracted, transmitted, or absorbed?

Q41. When light strikes a piece of white paper, the light is reflected in all directions. What do we call this scattering of light?

COMPARISON OF LIGHT WAVES WITH SOUND WAVES

There are two main differences between sound waves and light waves. The first difference is in velocity. Sound waves travel through air at the speed of approximately 1,100 feet per second; light waves travel through air and empty space at a speed of approximately 186,000 miles per second. The second difference is that sound is composed of longitudinal waves (alternate compressions and expansions of matter) and light is composed of transverse waves in an electromagnetic field.

Although both are forms of wave motion, sound requires a solid, liquid, or gaseous medium; whereas light travels through empty space. The denser the medium, the greater the speed of sound. The opposite is true of light. Light travels approximately one-third slower in water than in air. Sound travels through all substances, but light cannot pass through opaque materials.

Frequency affects both sound and light. A certain range of sound frequencies produces sensations that you can hear. A slow vibration (low frequency) in sound gives the sensation of a low note. A more rapid sound vibration (higher frequency) produces a higher note. Likewise, a certain range of light frequencies produces sensations that you can see. Violet light is produced at the high-frequency end of the light spectrum, while red light is produced at the low-frequency end of the light spectrum. A change in frequency of sound waves causes an audible sensation—a difference in pitch. A change in the frequency of a light wave causes a visual sensation—a difference in color.

For a comparison of light waves with sound waves, see table 1-2.

Table 1-2.—Comparison of Light Waves and Sound Waves

	SOUND WAVES	LIGHT WAVES
VELOCITY IN AIR	APPROXIMATELY 1,100 FEET PER SECOND	APPROXIMATELY 186,000 MILES PER SECOND
FORM	A FORM OF WAVE MOTION	A FORM OF WAVE MOTION
WAVE COMPOSITION	LONGITUDINAL	TRANSVERSE
TRANSMITTING MEDIUM	ALL SUBSTANCES	EMPTY SPACE AND ALL SUBSTANCES EXCEPT OPAQUE MATERIALS
RELATION OF TRANSMITTING MEDIUM VELOCITY TO VELOCITY	THE DENSER THE MEDIUM, THE GREATER THE SPEED	THE DENSER THE MEDIUM, THE SLOWER THE SPEED
SENSATIONS PRODUCED	HEARING	SEEING
VARIATIONS IN SENSATIONS PRODUCED	A LOW FREQUENCY CAUSES A LOW NOTE; A HIGH FREQUENCY, A HIGH NOTE	A LOW FREQUENCY CAUSES RED LIGHT; A HIGH FREQUENCY, VIOLET LIGHT

Q42. What three examples of electromagnetic energy are mentioned in the text?

Q43. What is the main difference between the bulk of the electromagnetic spectrum and the visual spectrum?

ELECTROMAGNETIC SPECTRUM

Light is one kind of electromagnetic energy. There are many other types, including heat energy and radio energy. The only difference between the various types of electromagnetic energy is the frequency of their waves (rate of vibration). The term SPECTRUM is used to designate the entire range of electromagnetic waves arranged in order of their frequencies. The VISIBLE SPECTRUM contains only those waves which stimulate the sense of sight. You, as a technician, might be expected to maintain equipment that uses electromagnetic waves within, above, and below the visible spectrum.

There are neither sharp dividing lines nor gaps in the ELECTROMAGNETIC SPECTRUM. Figure 1-22 illustrates how portions of the electromagnetic spectrum overlap. Notice that only a small portion of the electromagnetic spectrum contains visible waves, or light, which can be seen by the human eye.

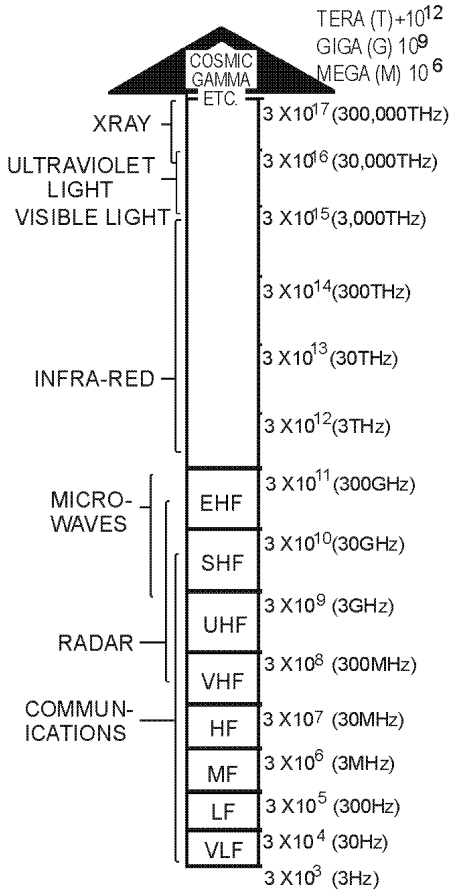


Figure 1-22.—Electromagnetic spectrum.

ELECTROMAGNETIC WAVES

In general, the same principles and properties of light waves apply to the communications electromagnetic waves you are about to study. The electromagnetic field is used to transfer energy (as communications) from point to point. We will introduce the basic ANTENNA as the propagation source of these electromagnetic waves.

THE BASIC ANTENNA

The study of antennas and electromagnetic wave propagation is essential to a complete understanding of radio communication, radar, loran, and other electronic systems. Figure 1-23 shows a simple radio communication system. In the illustration, the transmitter is an electronic device that generates radio-frequency energy. The energy travels through a transmission line (we will discuss this in chapter 3) to an antenna. The antenna converts the energy into radio waves that radiate into space from the antenna at the speed of light. The radio waves travel through the atmosphere or space until they are either reflected by an object or absorbed. If another antenna is placed in the path of the radio waves, it absorbs part of the waves and converts them to energy. This energy travels through another transmission line and is fed to a receiver. From this example, you can see that the requirements for a simple communications system are (1) transmitting equipment, (2) transmission line, (3) transmitting antenna, (4) medium, (5) receiving antenna, and (6) receiving equipment.

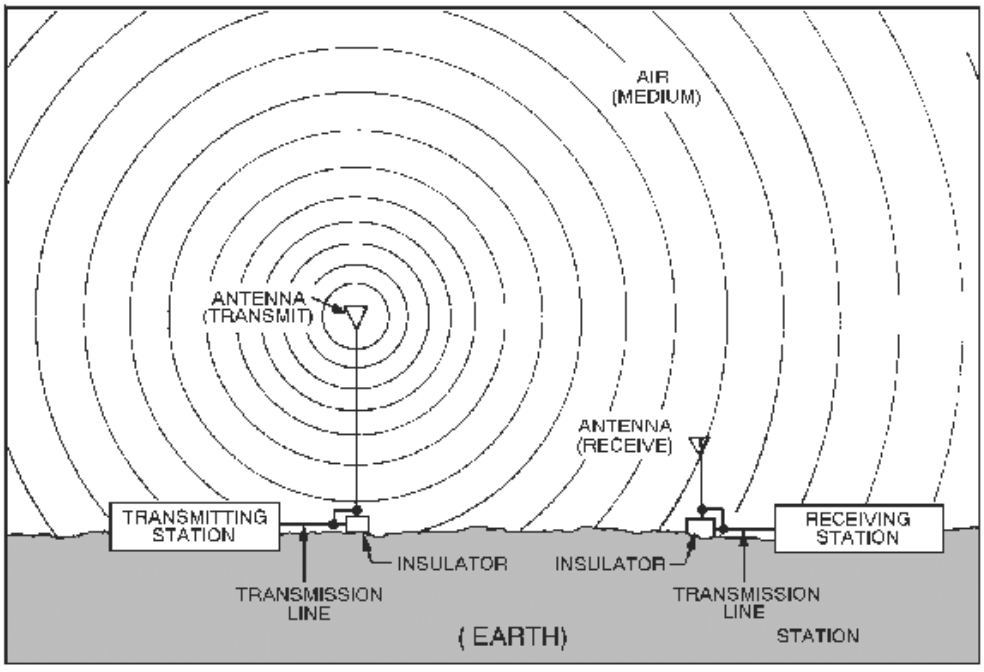


Figure 1-23.—Simple radio communication system.

An antenna is a conductor or a set of conductors used either to radiate electromagnetic energy into space or to collect this energy from space. Figure 1-24 shows an antenna. View A is a drawing of an actual antenna; view B is a cut-away view of the antenna; and view C is a simplified diagram of the antenna.

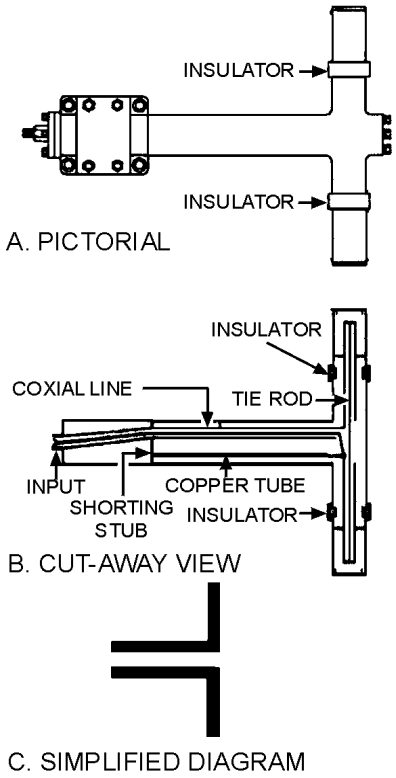


Figure 1-24.—Antenna.

COMPONENTS OF THE ELECTROMAGNETIC WAVE

An electromagnetic wave consists of two primary components—an ELECTRIC FIELD and a MAGNETIC FIELD. The electric field results from the force of voltage, and the magnetic field results from the flow of current.

Although electromagnetic fields that are radiated are commonly considered to be waves, under certain circumstances their behavior makes them appear to have some of the properties of particles. In general, however, it is easier to picture electromagnetic radiation in space as horizontal and vertical lines of force oriented at right angles to each other. These lines of force are made up of an electric field (E) and a magnetic field (H), which together make up the electromagnetic field in space.

The electric and magnetic fields radiated from an antenna form the electromagnetic field. This field is responsible for the transmission and reception of electromagnetic energy through free space. An antenna, however, is also part of the electrical circuit of a transmitter or a receiver and is equivalent to a circuit containing inductance, capacitance, and resistance. Therefore, the antenna can be expected to display definite voltage and current relationships with respect to a given input. A current through the antenna produces a magnetic field, and a charge on the antenna produces an electric field. These two fields combine to form the INDUCTION field. To help you gain a better understanding of antenna theory, we must review some basic electrical concepts. We will review voltage and its electric field, current and its magnetic field, and their relationship to propagation of electrical energy.

Q44. What are the two components (fields) that make up the electromagnetic wave?

Q45. What do we call a conductor (or set of conductors) that radiates electromagnetic energy into space?

Electric Field

Around every electrically charged object is a force field that can be detected and measured. This force field can cause electric charges to move in the field. When an object is charged electrically, there is either a greater or a smaller concentration of electrons than normal. Thus, a difference of potential exists between a charged object and an uncharged object. An electric field is, therefore, associated with a difference of potential, or a voltage.

This invisible field of force is commonly represented by lines that are drawn to show the paths along which the force acts. The lines representing the electric field are drawn in the direction that a single positive charge would normally move under the influence of that field. A large electric force is shown by a large concentration of lines; a weak force is indicated by a few lines.

When a capacitor is connected across a source of voltage, such as a battery, it is charged by a particular amount, depending on the voltage and the value of capacitance. (See figure 1-25.) Because of the emf (electromotive force) of the battery, negative charges flow to the lower plate, leaving the upper plate positively charged. Along with the growth of charge, the electric field is also building up. The flux lines are directed from the positive to the negative charges and at right angles to the plates. When the capacitor is fully charged, the voltage of the capacitor is equal to the voltage of the source and opposite in polarity. The charged capacitor stores the energy in the form of an electric field. It can be said, therefore, that an electric field indicates voltage.

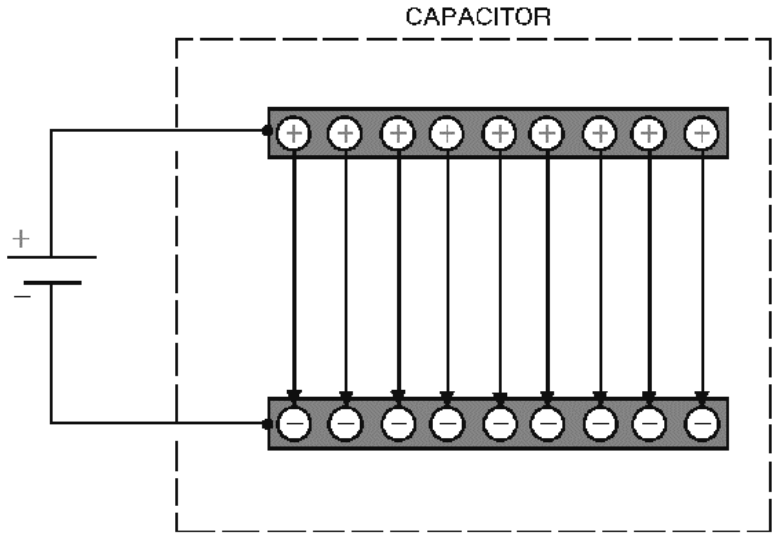


Figure 1-25.—Electric fields between plates.

If the two plates of the capacitor are spread farther apart, the electric field must curve to meet the plates at right angles (fig. 1-26). The straight lines in view A of figure 1-26 become arcs in view B, and approximately semicircles in view C, where the plates are in a straight line. Instead of flat metal plates, as in the capacitor, the two elements can take the form of metal rods or wires and form the basic antenna.

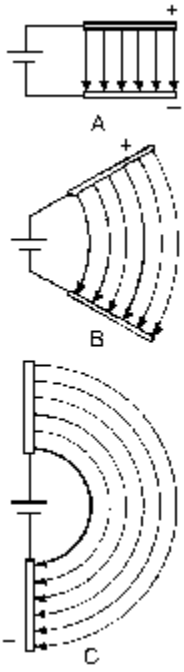


Figure 1-26.—Electric fields between plates at different angles.

In figure 1-27, two rods replace the plates of the capacitor, and the battery is replaced by an ac source generating a 60-hertz signal. On the positive alternation of the 60-hertz generator, the electric field extends from the positively charged rod to the negatively charged rod, as shown. On the negative alternation, the charge is reversed. The previous explanation of electrons moving from one plate to the other of the capacitor in figure 1-25 can also be applied to the rods in figure 1-27.

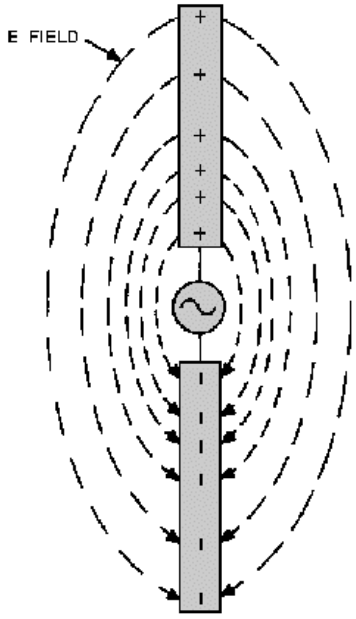


Figure 1-27.—Electric fields between elements.

The polarity of charges and the direction of the electric fields will reverse polarity and direction periodically at the frequency of the voltage source. The electric field will build up from zero to maximum in one direction and then collapse back to zero. Next, the field will build up to maximum in the opposite direction and then collapse back to zero. This complete reversal occurs during a single cycle of the source voltage. The HALF-WAVE DIPOLE ANTENNA (two separate rods in line as illustrated in figure 1-27) is the fundamental element normally used as a starting point of reference in any discussion concerning the radiation of electromagnetic energy into space. If rf energy from the ac generator (or transmitter) is supplied to the element of an antenna, the voltage across the antenna lags the current by 90 degrees. The antenna acts as if it were a capacitor.

Magnetic Field

When current flows through a conductor, a magnetic field is set up in the area surrounding the conductor. In fact, any moving electrical charge will create a magnetic field. The magnetic field is a region in space where a magnetic force can be detected and measured. There are two other fields involved—an INDUCTION FIELD, which exists close to the conductor carrying the current, and the RADIATION FIELD, which becomes detached from the current-carrying rod and travels through space.

To represent the magnetic field, lines of force are again used to illustrate the energy. Magnetic lines are not drawn between the rods, nor between high- and low-potential points, as the E lines that were discussed earlier. Magnetic lines are created by the flow of current rather than the force of voltage. The magnetic lines of force, therefore, are drawn at right angles to the direction of current flow.

The magnetic fields that are set up around two parallel rods, as shown in figure 1-28 view A, are in maximum opposition. Rod 1 contains a current flowing from the generator, while rod 2 contains a current flowing toward the generator. As a result, the direction of the magnetic field surrounding rod 1 is opposite the direction of the magnetic field surrounding rod 2. This will cause cancellation of part or all of both magnetic fields with a resultant decrease in radiation of the electromagnetic energy. View B illustrates the fact that if the far ends of rods 1 and 2 are separated from each other while the rods are still connected to the generator at the near ends, more space, and consequently less opposition, will occur between the magnetic fields of the two rods. View C illustrates the fact that placing the rods in line makes the currents through both rods flow in the same direction. Therefore, the two magnetic fields are in the same direction; thus, maximum electromagnetic radiation into space can be obtained.

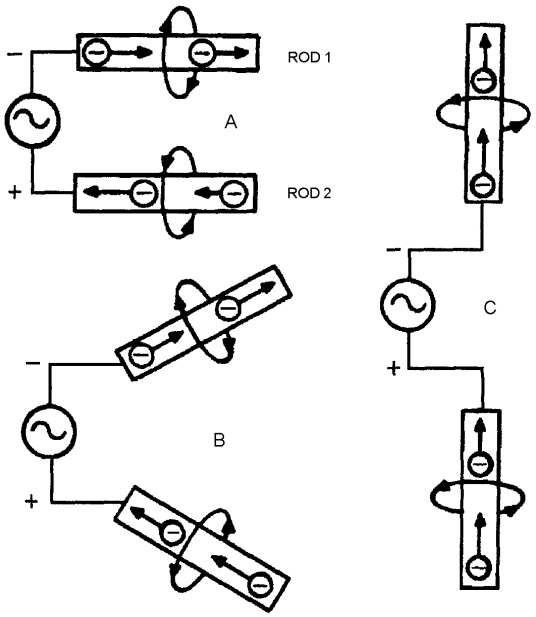


Figure 1-28.—Magnetic fields around elements.

Magnetic lines of force are indicated by the letter H and are called H lines. The direction of the magnetic lines may be determined by use of the left-hand rule for a conductor: If you grasp the conductor in your left hand with the thumb extended in the direction of the current flow, your fingers will point in the direction of the magnetic lines of force. In view C of figure 1-28, the direction of current flow is upward along both halves of the elements (conductors). The lines of magnetic force (flux) form concentric loops that are perpendicular to the direction of current flow. The arrowheads on the loops indicate the direction of the field. The left-hand rule is used to determine the direction of the magnetic field and is illustrated in figure 1-29. If the thumb of the left hand is extended in the direction of current flow and the fingers clenched, then the rough circles formed by the fingers indicate the direction of the magnetic field.

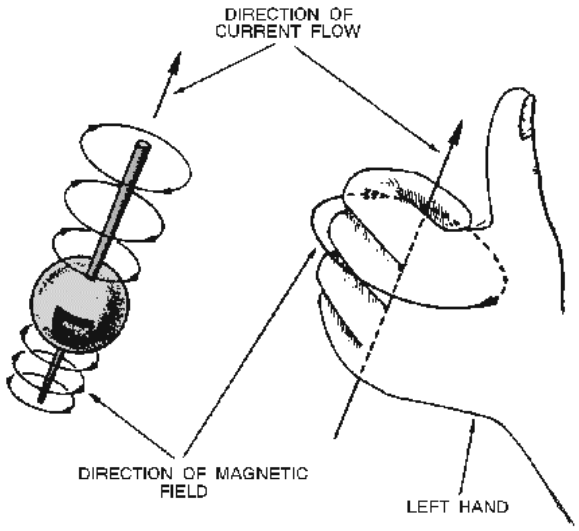


Figure 1-29.—Left-hand rule for conducting elements.

Q46. What do we call the field that is created between two rods when a voltage is applied to them?

Q47. When current flows through a conductor, a field is created around the conductor. What do we call this field?

Combined Electric and Magnetic Fields

The generator, shown in figure 1-30, provides the voltage, which creates an electric field, and current, which creates a magnetic field. This source voltage and current build up to maximum values in one direction during one half-cycle, and then build up to maximum values in the other direction during the next half-cycle. Both the electric and magnetic fields alternate from minimum through maximum values in synchronization with the changing voltage and current. The electric and magnetic fields reach their maximum intensity a quarter-cycle apart. These fields form the induction field. Since the current and voltage that produce these E and H fields are 90 degrees out of phase.

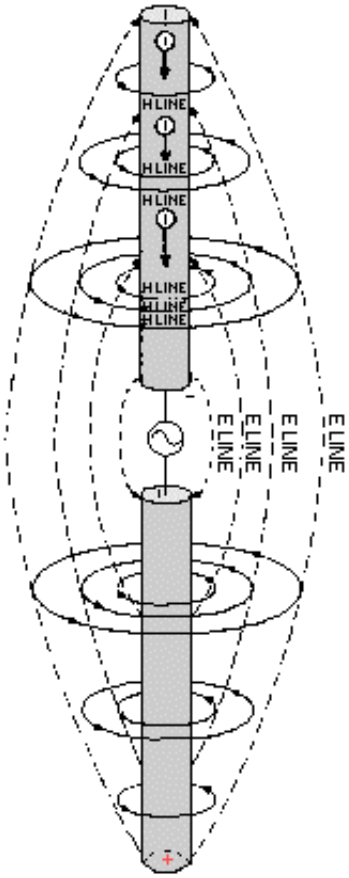


Figure 1-30.—Relationship of E-lines, and current flow.

Q48. An induction field is created around a conductor when current flows through it. What do we call the field that detaches itself from the conductor and travels through space?

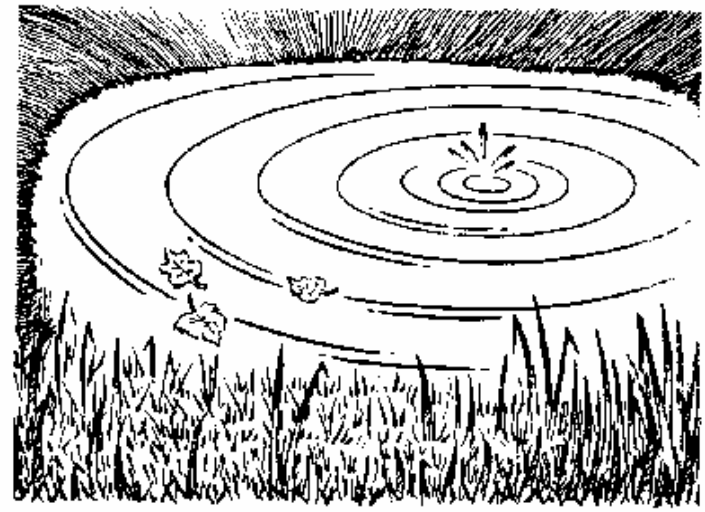
SUMMARY

Now that you have completed this chapter, let's review some of the new terms, concepts, and ideas you have learned. You should have a thorough understanding of these principles before moving on to chapter 2.

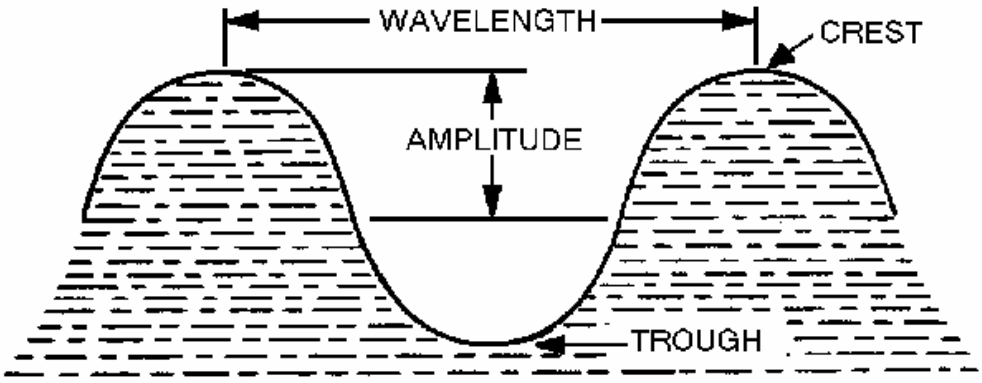
WAVE PROPAGATION is an invisible force that enables man to communicate over long distances. Wave transmission can take many forms, such as **LIGHT**, **SOUND**, and **RADIO**.

LIGHT is a form of wave motion that can be seen. Heat cannot normally be seen, but can be felt. Radio waves cannot be seen or felt.

WAVE MOTION can be seen in action by throwing a pebble into a pool of still water. The ripples that move toward the edge of the pool demonstrate the **PROPAGATION** theory.

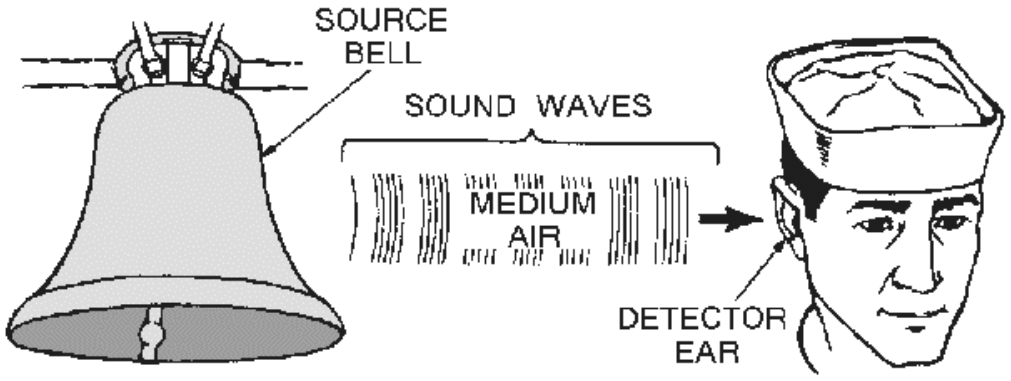


The **TRANSVERSE WAVE** is a type of wave motion. Radio, light, and heat waves are examples of transverse waves.



The **LONGITUDINAL WAVE** is another type of wave motion. The sound wave is the only example of a longitudinal wave given in this text.

SOURCE, MEDIUM, AND DETECTOR (RECEIVER) are the three requirements for all wave motion.



A **SOURCE** can be anything that emits or expends energy (waves).

The **MEDIUM** is the vehicle for carrying waves from one point to another. Water, air, metal, empty space, etc., are examples of a medium. Empty space is considered a medium for electro-magnetic waves but not a medium for sound waves.

The **SOUND DETECTOR** absorbs the waves emitted by the source. The human ear is an example of a detector.

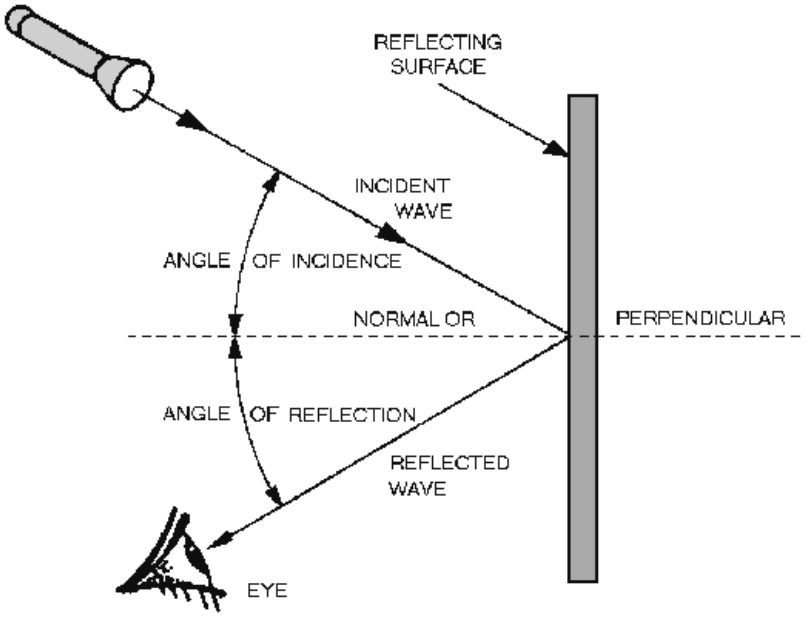
HERTZ, which is abbreviated **Hz**, is used in lieu of "cycle per second" when referring to radio frequencies.

VELOCITY OF PROPAGATION is the speed (or rate) at which the crest of a wave moves through a medium. Velocity can be calculated by using the formula:

$$V = \lambda f$$

Where v is velocity of propagation and is expressed in feet (meters) per second, λ is the wavelength in feet (meters), and f is the frequency in hertz.

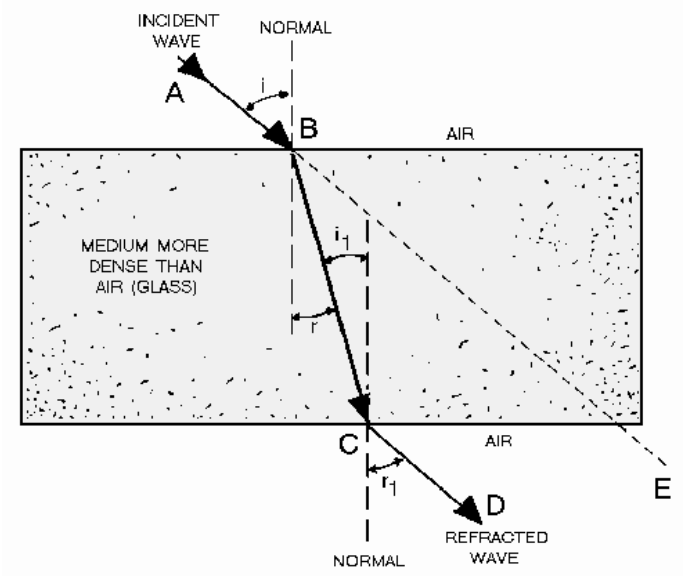
REFLECTION occurs when a wave strikes an object and bounces back (toward the source). The wave that moves from the source to the object is called the **INCIDENT WAVE**, and the wave that moves away from the object is called the **REFLECTED WAVE**.



The **LAW OF REFLECTION** states:

The angle of incidence is equal to the angle of reflection.

REFRACTION occurs when a wave traveling through two different mediums passes through the **BOUNDARY** of the mediums and bends toward or away from the **NORMAL**.



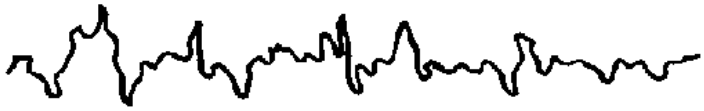
DIFFRACTION can account for the ability of the AM radio waves (due to their low frequency) to travel over a mountain, while FM and TV signals (due to their higher frequencies) are blocked.

The **DOPPLER EFFECT** is the apparent change in frequency of a source as it moves toward or away from a detector. It can affect the operation of equipment used to detect and measure wave energy.

SOUND can be audible to the human ear or it can be outside the hearing range.

NOISE AND TONES are the two general groups that broadly classify ALL sounds.

A MUSICAL NOTE



B NOISE

PITCH, INTENSITY, AND QUALITY are the three basic characteristics of sound. Pitch describes the frequency of sound. Intensity describes how much energy is transmitted. Quality enables us to distinguish one sound from another.

The **DENSITY** of a **MEDIUM**, **TEMPERATURE**, and **ATMOSPHERIC PRESSURE** affect the velocity of sound. If temperature, density, or pressure increases, the velocity of sound increases and vice versa.

ACOUSTICS is the science of sound and relates to the sense of hearing.

ECHO is an example of reflection. Sound echoes are used in sonar and depth finders to determine or measure the range of an object or the depth of the ocean bottom.

REVERBERATION is the multiple reflections of sound waves. The prolonged roar of thunder is caused by reverberations. With underwater sound equipment, reverberations of nearby objects may interfere with returning echoes from actual targets.

INTERFERENCE occurs when two waves move simultaneously through a medium. They can interfere constructively, destructively, or produce a resultant of zero.

RESONANCE occurs when an objects vibrates (or resonates) at its natural frequency. When different frequencies are produced inside a cavity, the sound from the cavity sounds louder at its resonant frequency than at all other frequencies.

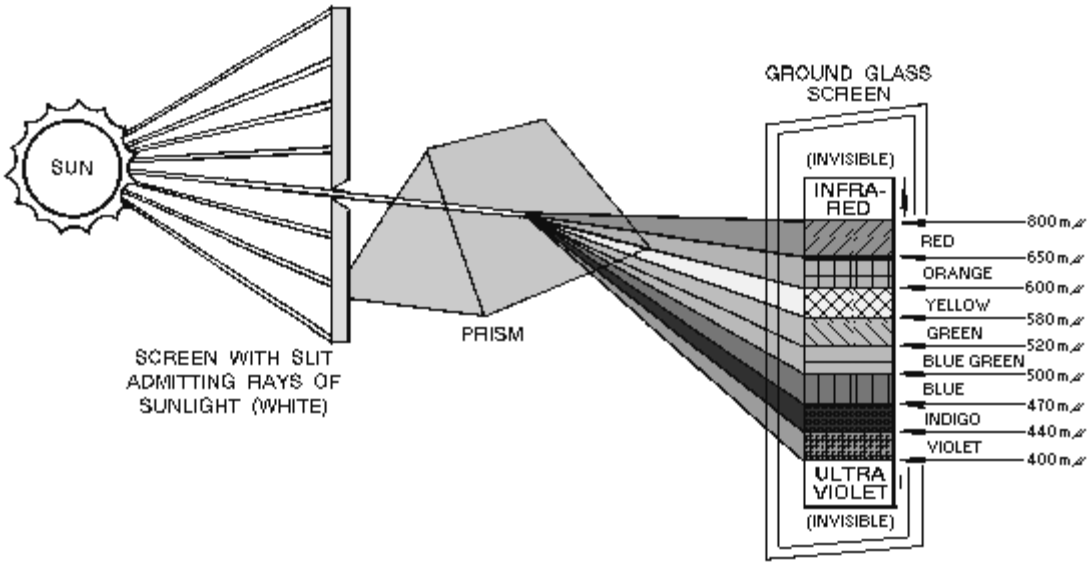
NOISE is any disturbance that distracts from or distorts the quality of sound.

A **PHOTON** is the smallest unit of radiant energy that makes up light waves and radio waves.

ANGSTROM (Å) units are used for measuring the wavelength of light. One angstrom = 1055^{-10} m.

The **VISIBLE SPECTRUM** contains all the colors between infrared and ultraviolet. **INFRA-RED** and **ULTRA-VIOLET** are invisible to the human eye.

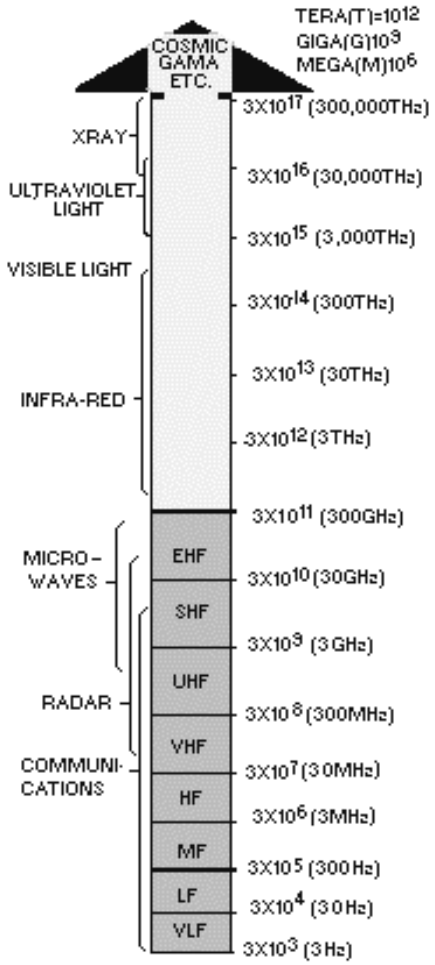
The **PRIMARY COLORS** of light are red, green, and blue. These primaries can be mixed to make any color between red and violet. If the three colors are mixed equally, they produce white light.



The **COMPLEMENTARY COLORS** of light are magenta, yellow, and cyan. They are produced by mixing any two of the primary colors together in overlapping beams.

The **SPEED OF LIGHT** in empty space is considered to be 186,000 miles per second (or 300,000,000 meters per second). This speed varies in different mediums, but the constant of 186,000 miles per second is always used as the speed of light.

The **ELECTROMAGNETIC SPECTRUM** is the complete range of electromagnetic frequencies from 3 kHz to beyond 300,000 THz. Audio frequencies (15 Hz-20 kHz) are not electromagnetic energy and are not included in the electromagnetic spectrum.



The **ELECTROMAGNETIC FIELD** consists of an **ELECTRIC FIELD** and a **MAGNETIC FIELD**. These fields are responsible for the transmission and reception of electromagnetic energy through free space.

ANSWERS TO QUESTIONS Q1. THROUGH Q48.

A1. Propagation means spreading out.

A2. A wave is a disturbance which moves through a medium.

A3. A means of transferring energy from one place to another.

A4. Sound waves, light waves, radio waves, heat waves, water waves.

A5. Transverse waves.

A6. Radio waves, light waves, and heat waves.

A7. A sound wave.

A8. A source, medium, and detector (receiver).

A9. A sequence of events, such as the positive and negative alternation of electrical current.

A10. The space occupied by one cycle of a radio wave at any given instant.

A11. The law of reflection states: The angle of incidence is equal to the angle of reflection.

A12. When the incident wave is nearly parallel with the surface.

A13. When the incident wave is perpendicular to the surface. Also a dull (or black) surface reflects very little regardless of the angle.

A14. The density of the two mediums, and the velocity of the waves.

A15. The Doppler effect.

A16. Sonics.

A17. No. The average human ear cannot hear all sounds in the infrasonic and ultrasonic regions.

A18. An amplifier.

A19. A source, medium, and detector (receiver).

A20. Noise and tones.

A21. Pitch, intensity, and quality.

A22. 20 Hz to 20 kHz.

A23. The amount of energy transmitted from a source.

A24. Quality.

A25. Velocity increases as density decreases and temperature increases.

A26. Acoustics.

A27. Echo.

A28. Reverberation.

A29. Resonance.

A30. Noise.

A31. Mechanical, electrical, and chemical.

A32. A photon.

A33. Angstrom unit.

A34. Red, green and blue.

CHAPTER 2

RADIO WAVE PROPAGATION

LEARNING OBJECTIVES

Upon completion of this unit, you should be able to:

1. State what the electromagnetic field is and what components make up the electromagnetic field.
2. State the difference between the induction field and the radiation field.
3. State what radio waves are.
4. List the components of a radio wave and define the terms cycle, frequency, harmonics, period, wavelength, and velocity as applied to radio wave propagation.
5. Compute the wavelength of radio waves.
6. State how radio waves are polarized, vertically and horizontally.
7. State what reflection, refraction, and diffraction are as applied to radio waves.
8. State what influence the Earth's atmosphere has on radio waves and list the different layers of the Earth's atmosphere.
9. Identify a ground wave, a sky wave, and state the effects of the ionosphere on the sky wave.
10. Identify the structure of the ionosphere.
11. Define density of layer, frequency, angle of incidence, skip distance, and skip zone.
12. Describe propagation paths.
13. Describe fading, multipath fading, and selective fading. Describe propagation paths.
14. State how transmission losses affect radio wave propagation.
15. State how electromagnetic interference, man-made/natural interference, and ionospheric disturbances affect radio wave propagation. State how transmission losses affect radio wave propagation.
16. Identify variations in the ionosphere.
17. Identify the maximum, optimum, and lowest usable frequencies of radio waves.
18. State what temperature inversion is, how frequency predictions are made, and how weather affects frequency.
19. State what tropospheric scatter is and how it affects radio wave propagation.

ELECTROMAGNETIC FIELDS

The way energy is propagated into free space is a source of great dispute among people concerned with it. Although many theories have been proposed, the following theory adequately explains the phenomena and has been widely accepted. There are two basic fields associated with every antenna; an INDUCTION FIELD and a RADIATION FIELD. The field associated with the energy stored in the antenna is the induction field. This field is said to provide no part in the transmission of electromagnetic energy through free space. However, without the presence of the induction field, there would be no energy radiated.

INDUCTION FIELD

Figure 2-1, a low-frequency generator connected to an antenna, will help you understand how the induction field is produced. Let's follow the generator through one cycle of operation.

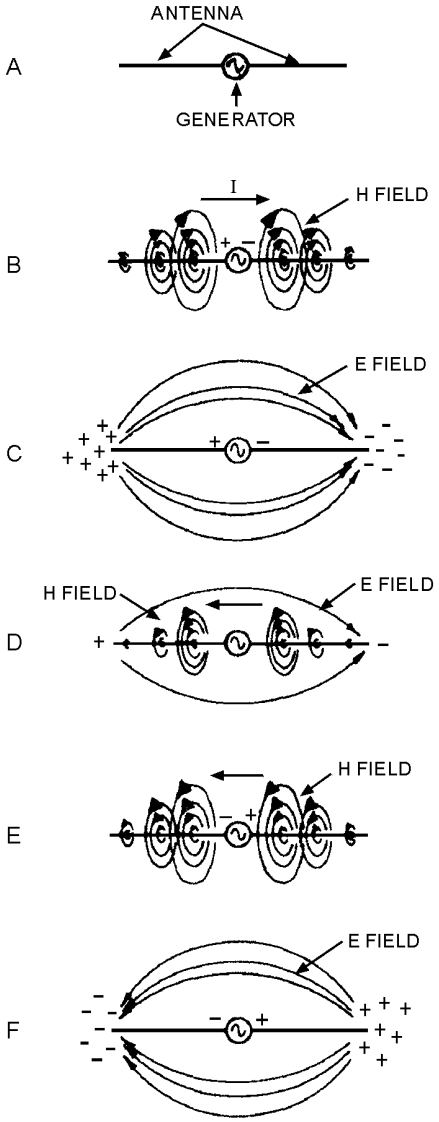


Figure 2-1.—Induction field about an antenna.

Initially, you can consider that the generator output is zero and that no fields exist about the antenna, as shown in view A. Now assume that the generator produces a slight potential and has the instantaneous polarity shown in view B. Because of this slight potential, the antenna capacitance acts as a short, allowing a large flow of current (I) through the antenna in the direction shown. This current flow, in turn, produces a large magnetic field about the antenna. Since the flow of current at each end of the antenna is minimum, the corresponding magnetic fields at each end of the antenna are also minimum. As time passes, charges, which oppose antenna current and produce an electrostatic field (E field), collect at each end of the antenna. Eventually, the antenna capacitance becomes fully charged and stops current flow through the antenna. Under this condition, the electrostatic field is maximum, and the magnetic field (H field) is fully collapsed, as shown in view C.

As the generator potential decreases back to zero, the potential of the antenna begins to discharge. During the discharging process, the electrostatic field collapses and the direction of current flow reverses, as shown in view D. When the current again begins to flow, an associated magnetic field is generated. Eventually, the electrostatic field completely collapses, the generator potential reverses, and current is maximum, as shown in view E. As charges collect at each end of the antenna, an electrostatic field is produced and current flow decreases. This causes the magnetic field to begin collapsing. The collapsing magnetic field produces more current flow, a greater accumulation of charge, and a greater electrostatic field. The antenna gradually reaches the condition shown in view F, where current is zero and the collected charges are maximum.

As the generator potential again decreases toward zero, the antenna begins to discharge and the electrostatic field begins to collapse. When the generator potential reaches zero, discharge current is maximum and the associated magnetic field is maximum. A brief time later, generator potential reverses, and the condition shown in view B recurs.

NOTE: The electric field (E field) and the electrostatic field (E field) are the same. They will be used interchangeably throughout this text.

The graph shown in figure 2-2 shows the relationship between the magnetic (H) field and the electric (E) field plotted against time. Note that the two fields are 90 degrees out of phase with each other. If you compare the graph in figure 2-2 with figure 2-1, you will notice that the two fields around the antenna are displaced 90 degrees from each other in space. (The H field exists in a plane perpendicular to the antenna. The E field exists in a plane parallel with the antenna, as shown in figure 2-1.)

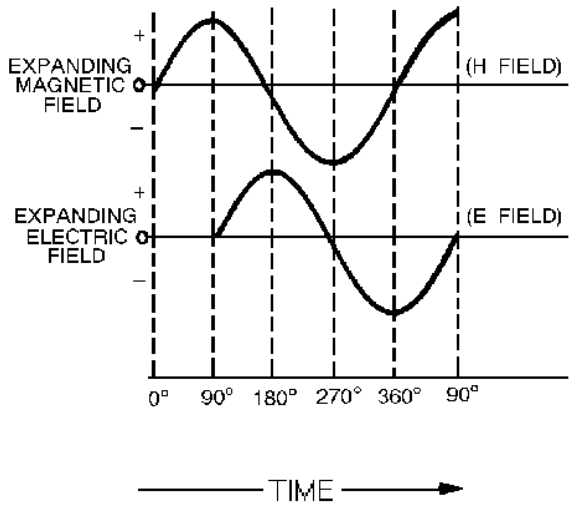


Figure 2-2.—Phase relationship of induction field components.

All the energy supplied to the induction field is returned to the antenna by the collapsing E and H fields. No energy from the induction field is radiated from the antenna. Therefore, the induction field is considered a local field and plays no part in the transmission of electromagnetic energy. The induction field represents only the stored energy in the antenna and is responsible only for the resonant effects that the antenna reflects to the generator.

RADIATION FIELDS

The E and H fields that are set up in the transfer of energy through space are known collectively as the radiation field. This radiation field is responsible for electromagnetic radiation from the antenna. The radiation field decreases as the distance from the antenna is increased. Because the decrease is linear, the radiation field reaches great distances from the antenna.

Let's look at a half-wave antenna to illustrate how this radiation actually takes place. Simply stated, a half-wave antenna is one that has an electrical length equal to half the wavelength of the signal being transmitted. Assume, for example, that a transmitter is operating at 30 megahertz. If a half-wave antenna is used with the transmitter, the antenna's electrical length would have to be at least 16 feet long. (The formula used to compute the electrical length of an antenna will be explained in chapter 4.) When power is delivered to the half-wave antenna, both an induction field and a radiation field are set up by the fluctuating energy. At the antenna, the intensities of these fields are proportional to the amount of power delivered to the antenna from a source such as a transmitter. At a short distance from the antenna and beyond, only the radiation field exists. This radiation field is made up of an electric component and a magnetic component at right angles to each other in space and varying together in intensity.

With a high-frequency generator (a transmitter) connected to the antenna, the induction field is produced as described in the previous section. However, the generator potential reverses before the electrostatic field has had time to collapse completely. The reversed generator potential neutralizes the remaining antenna charges, leaving a resultant E field in space.

Figure 2-3 is a simple picture of an E field detaching itself from an antenna. (The H field will not be considered, although it is present.) In view A the voltage is maximum and the electric field has maximum intensity. The lines of force begin at the end of the antenna that is positively charged and extend to the end of the antenna that is negatively charged. Note that the outer E lines are stretched away from the inner lines. This is because of the repelling force that takes place between lines of force in the same direction. As the voltage drops (view B), the separated charges come together, and the ends of the lines move toward the center of the antenna. But, since lines of force in the same direction repel each other, the centers of the lines are still being held out.

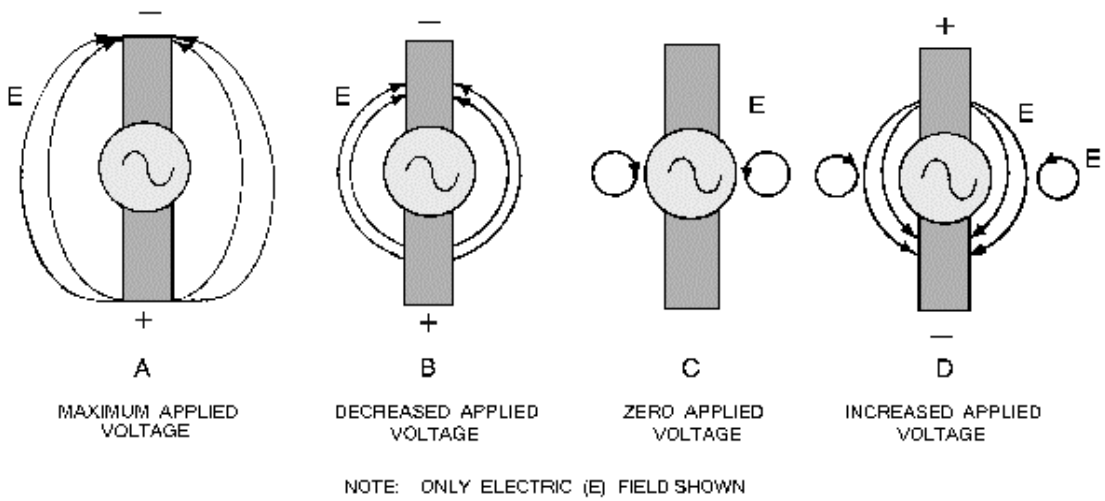


Figure 2-3.—Radiation from an antenna.

As the voltage approaches zero (view B), some of the lines collapse back into the antenna. At the same time, the ends of other lines begin to come together to form a complete loop. Notice the direction of these lines of force next to the antenna in view C. At this point the voltage on the antenna is zero. As the charge starts to build up in the opposite direction (view D), electric lines of force again begin at the positive end of the antenna and stretch to the negative end of the antenna. These lines of force, being in the same direction as the sides of the closed loops next to the antenna, repel the closed loops and force them out into space at the speed of light. As these loops travel through space, they generate a magnetic field in phase with them.

Since each successive E field is generated with a polarity that is opposite the preceding E field (that is, the lines of force are opposite), an oscillating electric field is produced along the path of travel. When an electric field oscillates, a magnetic field having an intensity that varies directly with that of the E field is produced. The variations in magnetic field intensity, in turn, produce another E field. Thus, the two varying fields sustain each other, resulting in electromagnetic wave propagation.

During this radiation process, the E and H fields are in phase in time but physically displaced 90 degrees in space. Thus, the varying magnetic field produces a varying electric field; and the varying electric field, in turn, sustains the varying magnetic field. Each field supports the other, and neither can be propagated by itself. Figure 2-4 shows a comparison between the induction field and the radiation field.

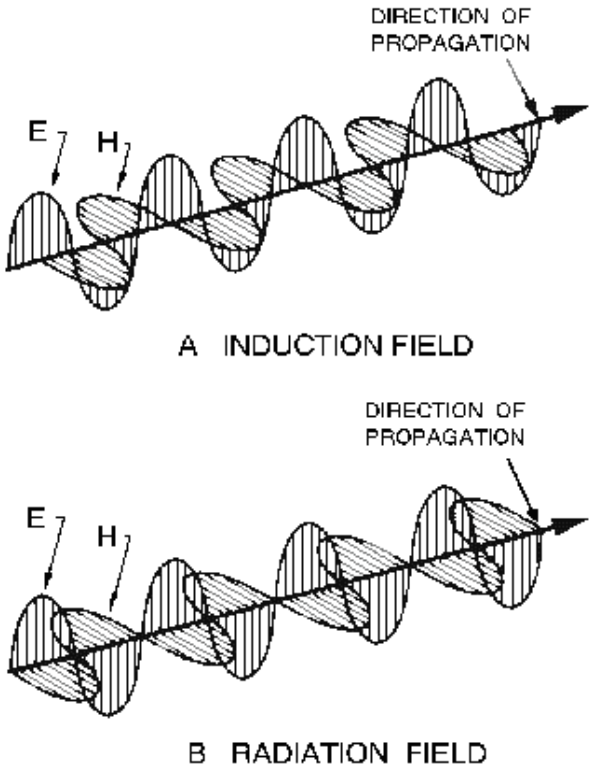


Figure 2-4.—E and H components of induction and radiation fields.

Q1. Which two composite fields (composed of E and H fields) are associated with every antenna?

Q2. What composite field (composed of E and H fields) is found stored in the antenna?

Q3. What composite field (composed of E and H fields) is propagated into free space?

RADIO WAVES

An energy wave generated by a transmitter is called a RADIO WAVE. The radio wave radiated into space by the transmitting antenna is a very complex form of energy containing both electric and magnetic fields. Because of this combination of fields, radio waves are also referred to as ELECTROMAGNETIC RADIATION.

This discussion will explain the Earth's atmosphere and its effect on radio waves. All the principles of wave motion that were discussed in chapter 1 also apply to radio waves.

NOTE: The term *radio wave* is not limited to communications equipment alone. The term applies to all equipment that generate signals in the form of electromagnetic energy.

COMPONENTS OF RADIO WAVES

The basic shape of the wave generated by a transmitter is that of a sine wave. The wave radiated out into space, however, may or may not retain the characteristics of the sine wave.

A sine wave can be one cycle or many cycles. Recall from chapter 1 that the number of cycles of a sine wave that are completed in 1 second is known as the *frequency* of the sine wave. For example, 60 cycles of ordinary house current occur each second, so house current is said to have a frequency of 60 cycles per second or 60 hertz.

The frequencies falling between 3000 hertz (3 kHz) and 300,000,000,000 hertz (300 GHz) are called RADIO FREQUENCIES (abbreviated rf) since they are commonly used in radio communications. This part of the radio frequency spectrum is divided into bands, each band being 10 times higher in frequency than the one immediately below it. This arrangement serves as a convenient way to remember the range of each band. The rf bands are shown in table 2-1. The usable radio-frequency range is roughly 10 kilohertz to 100 gigahertz.

Table 2-1.—Radio Frequency Bands

DESCRIPTION	ABBREVIATION	FREQUENCY
Very low	VLF	3 to 30 KHz
Low	LF	30 to 300 KHz
Medium	MF	300 to 3000 KHz
High	HF	3 to 30 MHz
Very high	VHF	30 to 300 MHz
Ultrahigh	UHF	300 to 3000 MHz
Super high	SHF	3 to 30 GHz
Extremely high	EHF	30 to 300 GHz

Any frequency that is a whole number multiple of a smaller basic frequency is known as a HARMONIC of that basic frequency. The basic frequency itself is called the first harmonic or, more commonly, the FUNDAMENTAL FREQUENCY. A frequency that is twice as great as the fundamental frequency is called the second harmonic; a frequency three times as great is the third harmonic; and so on. For example:

First harmonic (Fundamental frequency)	3000 kHz
Second harmonic	6000 kHz
Third harmonic	9000 kHz

The PERIOD of a radio wave is simply the amount of time required for the completion of one full cycle. If a sine wave has a frequency of 2 hertz, each cycle has a duration, or period, of one-half second. If the frequency is 10 hertz, the period of each cycle is one-tenth of a second. Since the frequency of a radio wave is the number of cycles that are completed in one second, you should be able to see that as the frequency of a radio wave increases, its period decreases.

A wavelength is the space occupied by one full cycle of a radio wave at any given instant. Wavelengths are expressed in meters (1 meter is equal to 3.28 feet). You need to have a good understanding of frequency and wavelength to be able to select the proper antenna(s) for use in successful

communications. The relationship between frequency, wavelength, and antennas will be discussed in chapter 4 of this module.

The velocity (or speed) of a radio wave radiated into free space by a transmitting antenna is equal to the speed of light—186,000 miles per second or 300,000,000 meters per second. Because of various factors, such as barometric pressure, humidity, molecular content, etc., radio waves travel inside the Earth's atmosphere at a speed slightly less than the speed of light. Normally, in discussions of the velocity of radio waves, the velocity referred to is the speed at which radio waves travel in free space.

The frequency of a radio wave has nothing to do with its velocity. A 5-megahertz wave travels through space at the same velocity as a 10-megahertz wave. However, the velocity of radio waves is an important factor in making wavelength-to-frequency conversions, the subject of our next discussion.

Q4. What is the term used to describe the basic frequency of a radio wave?

Q5. What is the term used to describe a whole number multiple of the basic frequency of a radio wave?

WAVELENGTH-TO-FREQUENCY CONVERSIONS

Radio waves are often referred to by their wavelength in meters rather than by frequency. For example, most people have heard commercial radio stations make announcements similar to the following: "Station WXYZ operating on 240 meters..." To tune receiving equipment that is calibrated by frequency to such a station, you must first convert the designated wavelength to its equivalent frequency.

As discussed earlier, a radio wave travels 300,000,000 meters a second (speed of light); therefore, a radio wave of 1 hertz would have traveled a distance (or wavelength) of 300,000,000 meters. Obviously then, if the frequency of the wave is increased to 2 hertz, the wavelength will be cut in half to 150,000,000 meters. This illustrates the principle that the HIGHER THE FREQUENCY, the SHORTER THE WAVELENGTH.

Wavelength-to-frequency conversions of radio waves are really quite simple because wavelength and frequency are reciprocals: Either one divided into the velocity of a radio wave yields the other. Remember, the formula for wavelength is:

$$\lambda = \frac{v}{f} \quad \text{or} \quad f = \frac{v}{\lambda}$$

Where:

λ = wavelength in meters

v = velocity of radio wave (speed of light)

f = frequency of radio wave (in Hz, kHz or Mhz)

The wavelength in meters divided into 300,000,000 yields the frequency of a radio wave in hertz. Likewise, the wavelength divided into 300,000 yields the frequency of a radio wave in kilohertz, and the wavelength divided into 300 yields the frequency in megahertz.

Now, let us apply the formula to determine the frequency to which the receiving equipment must be tuned to receive station WXYZ operating on 240 meters. Radio wave frequencies are normally expressed in kilohertz or megahertz.

To find the frequency in hertz, use the formula:

$$f = \frac{v}{\lambda}$$

Given:

$v = 300{,}000{,}000$ meters per second

$\lambda = 240$ meters

Solution:

$$f = \frac{300{,}000{,}000 \text{ meters per second}}{240 \text{ meters}}$$

$f = 1{,}250{,}000$ Hz

To find the frequency in kilohertz, use the formula:

$$f_{[kHz]} = \frac{300{,}000}{\lambda}$$

Given:

$\lambda = 240$ meters

Solution:

$$f_{[kHz]} = \frac{300{,}000}{240 \text{ meters}}$$

$f = 1250$kHz

To find the frequency in megahertz, use the formula:

$$f_{[MHz]} = \frac{300}{\lambda}$$

Given:

$\lambda = 240$ meters

Solution:

$$f_{[MHz]} = \frac{300}{240 \text{ meters}}$$

$f = 1.25$MHz

Q6. It is known that WWV operates on a frequency of 10 megahertz. What is the wavelength of WWV?

Q7. A station is known to operate at 60-meters. What is the frequency of the unknown station?

POLARIZATION

For maximum absorption of energy from the electromagnetic fields, the receiving antenna must be located in the plane of polarization. This places the conductor of the antenna at right angles to the magnetic lines of force moving through the antenna and parallel to the electric lines, causing maximum induction.

Normally, the plane of polarization of a radio wave is the plane in which the E field propagates with respect to the Earth. If the E field component of the radiated wave travels in a plane perpendicular to the Earth's surface (vertical), the radiation is said to be VERTICALLY POLARIZED, as shown in figure 2-5, view A. If the E field propagates in a plane parallel to the Earth's surface (horizontal), the radiation is said to be HORIZONTALLY POLARIZED, as shown in view B.

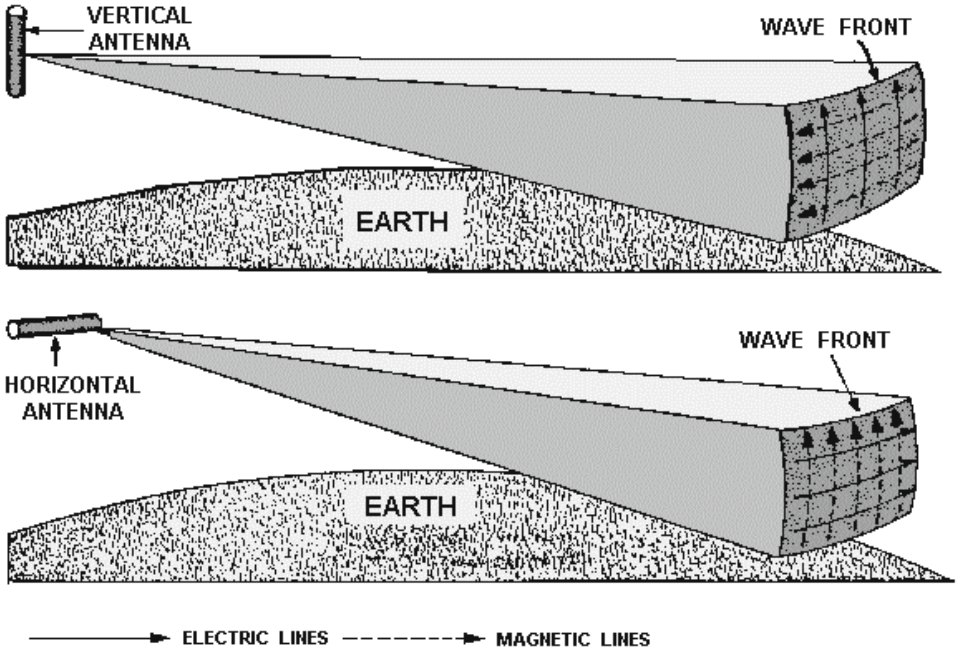


Figure 2-5.—Vertical and horizontal polarization.

The position of the antenna in space is important because it affects the polarization of the electromagnetic wave. When the transmitting antenna is close to the ground, vertically polarized waves cause a greater signal strength along the Earth's surface. On the other hand, antennas high above the ground should be horizontally polarized to get the greatest possible signal strength to the Earth's surface. Vertically and horizontally polarized antennas will be discussed in more detail in chapter 4.

The radiated energy from an antenna is in the form of an expanding sphere. Any small section of this sphere is perpendicular to the direction the energy travels and is called a WAVEFRONT. All energy on a wavefront is in phase. Usually all points on the wavefront are at equal distances from the antenna. The farther the wavefront is from the antenna, the less spherical the wave appears. At a considerable distance the wavefront can be considered as a plane surface at a right angle to the direction of propagation.

If you know the directions of the E and H components, you can use the "right-hand rule" (see figure 2-6) to determine the direction of wave propagation. This rule states that if the thumb, forefinger, and middle finger of the right hand are extended so they are mutually perpendicular, the middle finger will point in the direction of wave propagation if the thumb points in the direction of the E field and the forefinger points in the direction of the H field. Since both the E and H fields reverse directions simultaneously, propagation of a particular wavefront is always in the same direction (away from the antenna).

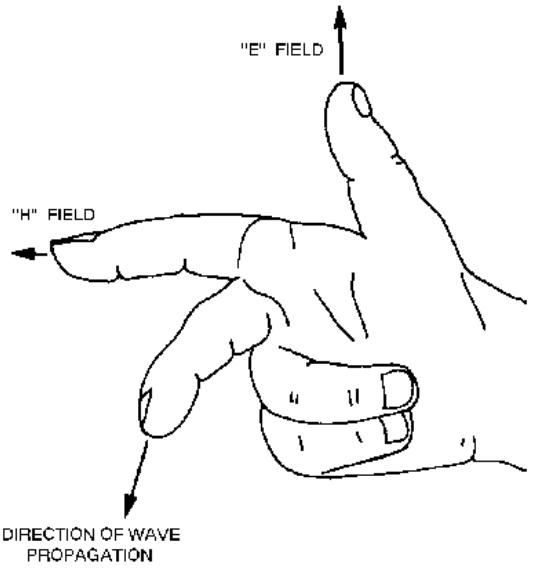


Figure 2-6.—Right-hand rule for propagation.

Q8. If a transmitting antenna is placed close to the ground, how should the antenna be polarized to give the greatest signal strength?

Q9. In the right-hand rule for propagation, the thumb points in the direction of the E field and the forefinger points in the direction of the H field. In what direction does the middle finger point?

ATMOSPHERIC PROPAGATION

Within the atmosphere, radio waves can be reflected, refracted, and diffracted like light and heat waves.

Reflection

Radio waves may be reflected from various substances or objects they meet during travel between the transmitting and receiving sites. The amount of reflection depends on the reflecting material. Smooth metal surfaces of good electrical conductivity are efficient reflectors of radio waves. The surface of the Earth itself is a fairly good reflector. The radio wave is not reflected from a single point on the reflector but rather from an area on its surface. The size of the area required for reflection to take place depends on the wavelength of the radio wave and the angle at which the wave strikes the reflecting substance.

When radio waves are reflected from flat surfaces, a phase shift in the alternations of the wave occurs. Figure 2-7 shows two radio waves being reflected from the Earth's surface. Notice that the positive and negative alternations of radio waves (A) and (B) are in phase with each other in their paths toward the Earth's surface. After reflection takes place, however, the waves are approximately 180 degrees out of phase from their initial relationship. The amount of phase shift that occurs is not constant.

It depends on the polarization of the wave and the angle at which the wave strikes the reflecting surface. Radio waves that keep their phase relationships after reflection normally produce a stronger signal at the receiving site. Those that are received out of phase produce a weak or fading signal. The shifting in the phase relationships of reflected radio waves is one of the major reasons for fading. Fading will be discussed in more detail later in this chapter.

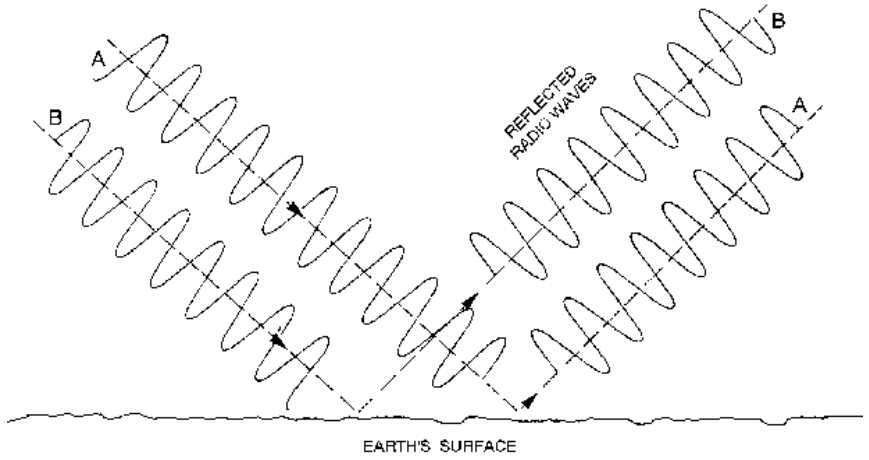


Figure 2-7.—Phase shift of reflected radio waves.

Refraction

Another phenomenon common to most radio waves is the bending of the waves as they move from one medium into another in which the velocity of propagation is different. This bending of the waves is called refraction. For example, suppose you are driving down a smoothly paved road at a constant speed and suddenly one wheel goes off onto the soft shoulder. The car tends to veer off to one side. The change of medium, from hard surface to soft shoulder, causes a change in speed or velocity. The tendency is for the car to change direction. This same principle applies to radio waves as changes occur in the medium through which they are passing. As an example, the radio wave shown in figure 2-8 is traveling through the Earth's atmosphere at a constant speed. As the wave enters the dense layer of electrically charged ions, the part of the wave that enters the new medium first travels faster than the parts of the wave that have not yet entered the new medium. This abrupt increase in velocity of the upper part of the wave causes the wave to bend back toward the Earth. This bending, or change of direction, is always toward the medium that has the lower velocity of propagation.

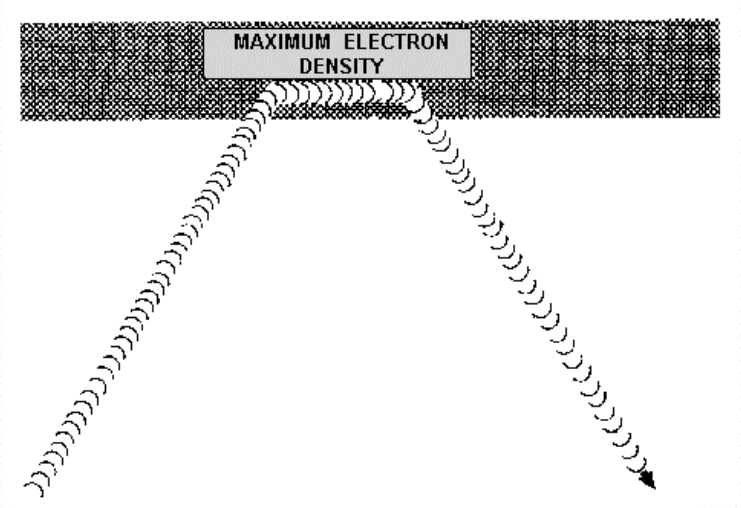


Figure 2-8.—Radio wave refraction.

Radio waves passing through the atmosphere are affected by certain factors, such as temperature, pressure, humidity, and density. These factors can cause the radio waves to be refracted. This effect will be discussed in greater detail later in this chapter.

Diffraction

A radio wave that meets an obstacle has a natural tendency to bend around the obstacle as illustrated in figure 2-9. The bending, called diffraction, results in a change of direction of part of the wave energy from the normal line-of-sight path. This change makes it possible to receive energy around the edges of an obstacle as shown in view A or at some distances below the highest point of an obstruction, as shown in view B. Although diffracted rf energy usually is weak, it can still be detected by a suitable receiver. The principal effect of diffraction extends the radio range beyond the visible horizon. In certain cases, by using high power and very low frequencies, radio waves can be made to encircle the Earth by diffraction.

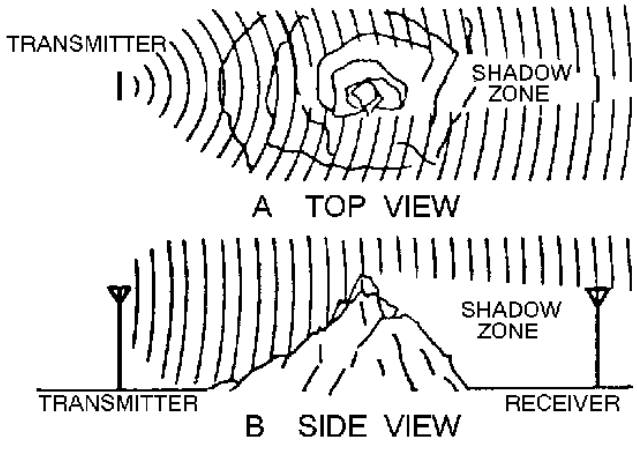


Figure 2-9.—Diffraction around an object.

Q10. What is one of the major reasons for the fading of radio waves which have been reflected from a surface?

THE EFFECT OF THE EARTH'S ATMOSPHERE ON RADIO WAVES

This discussion of electromagnetic wave propagation is concerned mainly with the properties and effects of the medium located between the transmitting antenna and the receiving antenna. While radio waves traveling in free space have little outside influence affecting them, radio waves traveling within the Earth's atmosphere are affected by varying conditions. The influence exerted on radio waves by the Earth's atmosphere adds many new factors to complicate what at first seems to be a relatively simple problem. These complications are because of a lack of uniformity within the Earth's atmosphere. Atmospheric conditions vary with changes in height, geographical location, and even with changes in time (day, night, season, year). A knowledge of the composition of the Earth's atmosphere is extremely important for understanding wave propagation.

The Earth's atmosphere is divided into three separate regions, or layers. They are the TROPOSPHERE, the STRATOSPHERE, and the IONOSPHERE. The layers of the atmosphere are illustrated in figure 2-10.

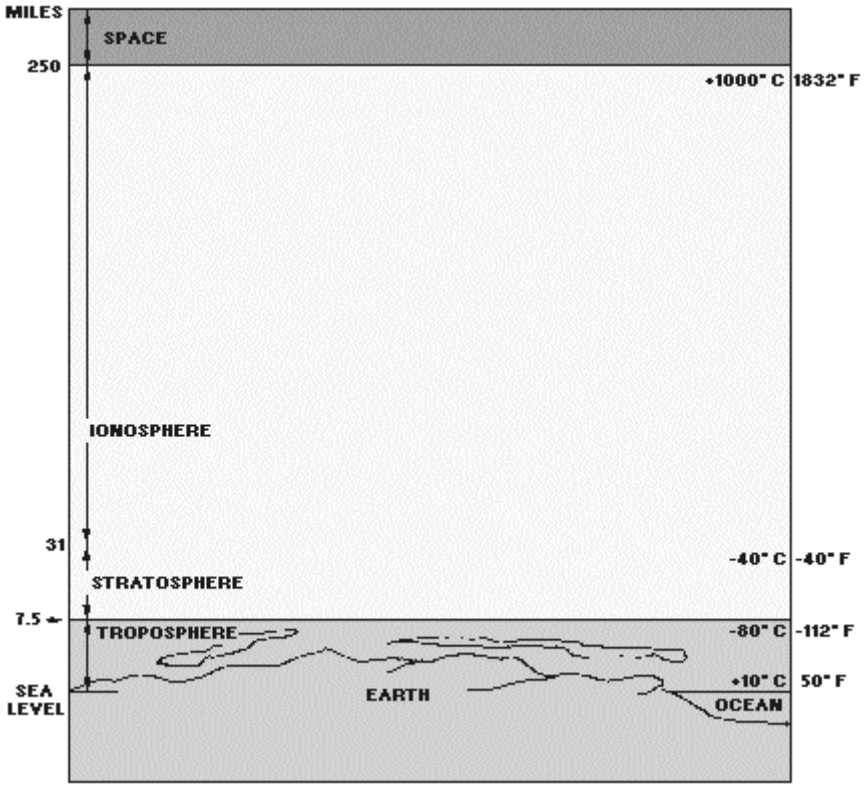


Figure 2-10.—Layers of the earth's atmosphere.

TROPOSPHERE

The troposphere is the portion of the Earth's atmosphere that extends from the surface of the Earth to a height of about 3.7 miles (6 km) at the North Pole or the South Pole and 11.2 miles (18 km) at the

equator. Virtually all weather phenomena take place in the troposphere. The temperature in this region decreases rapidly with altitude, clouds form, and there may be much turbulence because of variations in temperature, density, and pressure. These conditions have a great effect on the propagation of radio waves, which will be explained later in this chapter.

STRATOSPHERE

The stratosphere is located between the troposphere and the ionosphere. The temperature throughout this region is considered to be almost constant and there is little water vapor present. The stratosphere has relatively little effect on radio waves because it is a relatively calm region with little or no temperature changes.

IONOSPHERE

The ionosphere extends upward from about 31.1 miles (50 km) to a height of about 250 miles (402 km). It contains four cloud-like layers of electrically charged ions, which enable radio waves to be propagated to great distances around the Earth. This is the most important region of the atmosphere for long distance point-to-point communications. This region will be discussed in detail a little later in this chapter.

Q11. What are the three layers of the atmosphere?

Q12. Which layer of the atmosphere has relatively little effect on radio waves?

RADIO WAVE TRANSMISSION

There are two principal ways in which electromagnetic (radio) energy travels from a transmitting antenna to a receiving antenna. One way is by GROUND WAVES and the other is by SKY WAVES. Ground waves are radio waves that travel near the surface of the Earth (surface and space waves). Sky waves are radio waves that are reflected back to Earth from the ionosphere. (See figure 2-11.)

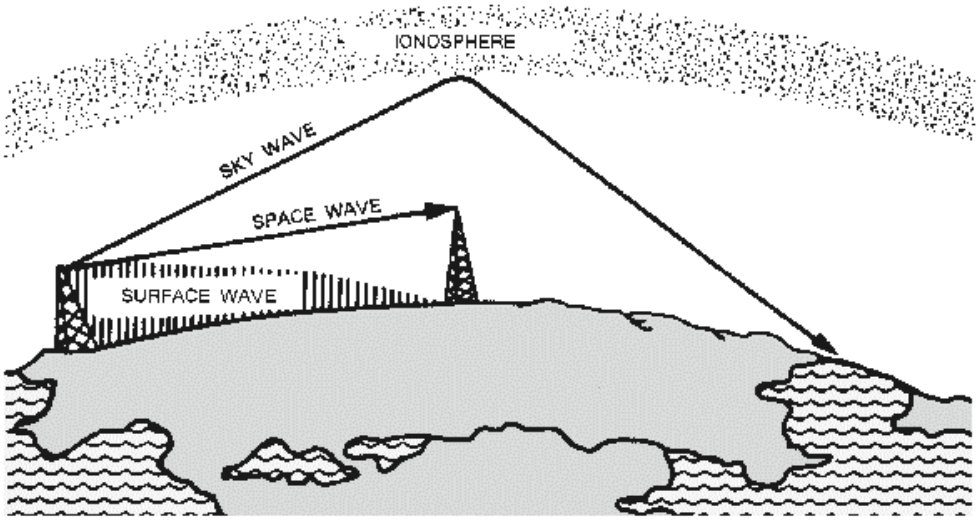


Figure 2-11.—Ground waves and sky waves.

Ground Waves

The ground wave is actually composed of two separate component waves. These are known as the SURFACE WAVE and the SPACE WAVE (fig. 2-11). The determining factor in whether a ground wave component is classified as a space wave or a surface wave is simple. A surface wave travels along the surface of the Earth. A space wave travels over the surface.

SURFACE WAVE.—The surface wave reaches the receiving site by traveling along the surface of the ground as shown in figure 2-12. A surface wave can follow the contours of the Earth because of the process of diffraction. When a surface wave meets an object and the dimensions of the object do not exceed its wavelength, the wave tends to curve or bend around the object. The smaller the object, the more pronounced the diffractive action will be.

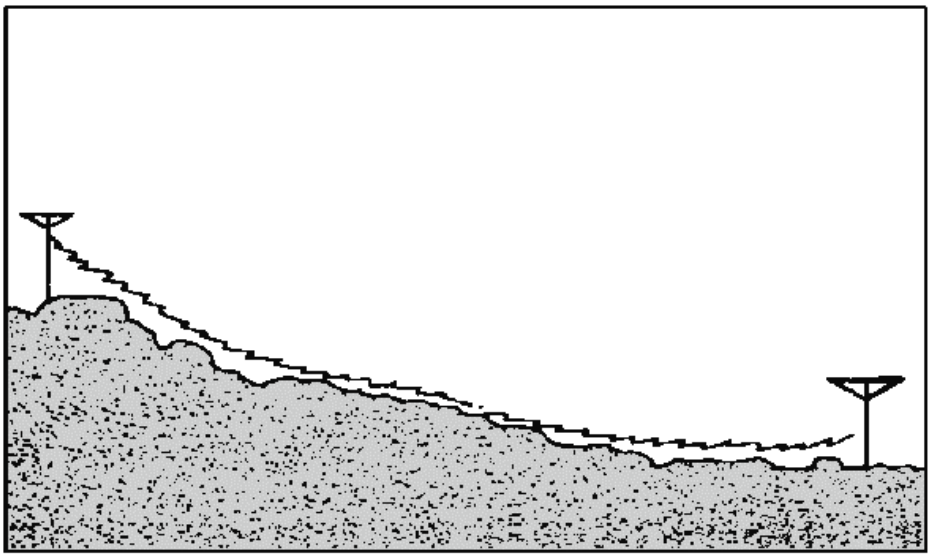


Figure 2-12.—Surface wave propagation.

As a surface wave passes over the ground, the wave induces a voltage in the Earth. The induced voltage takes energy away from the surface wave, thereby weakening, or attenuating, the wave as it moves away from the transmitting antenna. To reduce the attenuation, the amount of induced voltage must be reduced. This is done by using vertically polarized waves that minimize the extent to which the electric field of the wave is in contact with the Earth. When a surface wave is horizontally polarized, the electric field of the wave is parallel with the surface of the Earth and, therefore, is constantly in contact with it. The wave is then completely attenuated within a short distance from the transmitting site. On the other hand, when the surface wave is vertically polarized, the electric field is vertical to the Earth and merely dips into and out of the Earth's surface. For this reason, vertical polarization is vastly superior to horizontal polarization for surface wave propagation.

The attenuation that a surface wave undergoes because of induced voltage also depends on the electrical properties of the terrain over which the wave travels. The best type of surface is one that has good electrical conductivity. The better the conductivity, the less the attenuation. Table 2-2 gives the relative conductivity of various surfaces of the Earth.

Table 2-2.—Surface Conductivity

SURFACE	RELATIVE CONDUCTIVITY
Sea water	Good
Flat, loamy soil	Fair
Large bodies of fresh water	Fair
Rocky terrain	Poor
Desert	Poor
Jungle	Unusable

Another major factor in the attenuation of surface waves is frequency. Recall from earlier discussions on wavelength that the higher the frequency of a radio wave, the shorter its wavelength will be. These high frequencies, with their shorter wavelengths, are not normally diffracted but are absorbed by the Earth at points relatively close to the transmitting site. You can assume, therefore, that as the frequency of a surface wave is increased, the more rapidly the surface wave will be absorbed, or attenuated, by the Earth. Because of this loss by attenuation, the surface wave is impractical for long-distance transmissions at frequencies above 2 megahertz. On the other hand, when the frequency of a surface wave is low enough to have a very long wavelength, the Earth appears to be very small, and diffraction is sufficient for propagation well beyond the horizon. In fact, by lowering the transmitting frequency into the very low frequency (vlf) range and using very high-powered transmitters, the surface wave can be propagated great distances. The Navy's extremely high-powered vlf transmitters are actually capable of transmitting surface wave signals around the Earth and can provide coverage to naval units operating anywhere at sea.

SPACE WAVE.—The space wave follows two distinct paths from the transmitting antenna to the receiving antenna—one through the air directly to the receiving antenna, the other reflected from the ground to the receiving antenna. This is illustrated in figure 2-13. The primary path of the space wave is directly from the transmitting antenna to the receiving antenna. So, the receiving antenna must be located within the radio horizon of the transmitting antenna. Because space waves are refracted slightly, even when propagated through the troposphere, the radio horizon is actually about one-third farther than the line-of-sight or natural horizon.

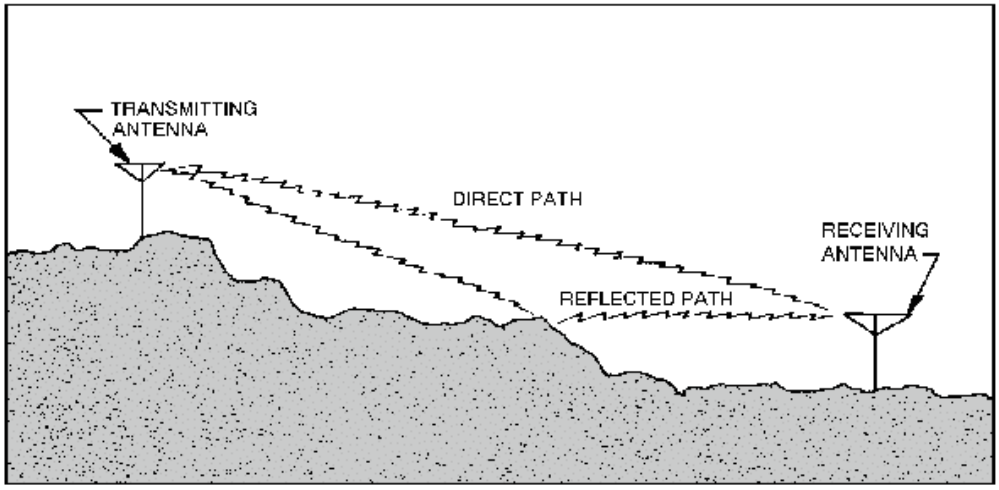


Figure 2-13.—Space wave propagation.

Although space waves suffer little ground attenuation, they nevertheless are susceptible to fading. This is because space waves actually follow two paths of different lengths (direct path and ground reflected path) to the receiving site and, therefore, may arrive in or out of phase. If these two component waves are received in phase, the result is a reinforced or stronger signal. Likewise, if they are received out of phase, they tend to cancel one another, which results in a weak or fading signal.

Q13. What is the determining factor in classifying whether a radio wave is a ground wave or a space wave?

Q14. What is the best type of surface or terrain to use for radio wave transmission?

Q15. What is the primary difference between the radio horizon and the natural horizon?

Q16. What three factors must be considered in the transmission of a surface wave to reduce attenuation?

Sky Wave

The sky wave, often called the ionospheric wave, is radiated in an upward direction and returned to Earth at some distant location because of refraction from the ionosphere. This form of propagation is relatively unaffected by the Earth's surface and can propagate signals over great distances. Usually the high frequency (hf) band is used for sky wave propagation. The following in-depth study of the ionosphere and its effect on sky waves will help you to better understand the nature of sky wave propagation.

STRUCTURE OF THE IONOSPHERE

As we stated earlier, the ionosphere is the region of the atmosphere that extends from about 30 miles above the surface of the Earth to about 250 miles. It is appropriately named the ionosphere because it consists of several layers of electrically charged gas atoms called ions. The ions are formed by a process called ionization.

Ionization

Ionization occurs when high energy ultraviolet light waves from the sun enter the ionospheric region of the atmosphere, strike a gas atom, and literally knock an electron free from its parent atom. A normal atom is electrically neutral since it contains both a positive proton in its nucleus and a negative orbiting electron. When the negative electron is knocked free from the atom, the atom becomes positively charged (called a positive ion) and remains in space along with the free electron, which is negatively charged. This process of upsetting electrical neutrality is known as IONIZATION.

The free negative electrons subsequently absorb part of the ultraviolet energy, which initially freed them from their atoms. As the ultraviolet light wave continues to produce positive ions and negative electrons, its intensity decreases because of the absorption of energy by the free electrons, and an ionized layer is formed. The rate at which ionization occurs depends on the density of atoms in the atmosphere and the intensity of the ultraviolet light wave, which varies with the activity of the sun.

Since the atmosphere is bombarded by ultraviolet light waves of different frequencies, several ionized layers are formed at different altitudes. Lower frequency ultraviolet waves penetrate the atmosphere the least; therefore, they produce ionized layers at the higher altitudes. Conversely, ultraviolet waves of higher frequencies penetrate deeper and produce layers at the lower altitudes.

An important factor in determining the density of ionized layers is the elevation angle of the sun, which changes frequently. For this reason, the height and thickness of the ionized layers vary, depending on the time of day and even the season of the year.

Recombination

Recall that the process of ionization involves ultraviolet light waves knocking electrons free from their atoms. A reverse process called RECOMBINATION occurs when the free electrons and positive ions collide with each other. Since these collisions are inevitable, the positive ions return to their original neutral atom state.

The recombination process also depends on the time of day. Between the hours of early morning and late afternoon, the rate of ionization exceeds the rate of recombination. During this period, the ionized layers reach their greatest density and exert maximum influence on radio waves. During the late afternoon and early evening hours, however, the rate of recombination exceeds the rate of ionization, and the density of the ionized layers begins to decrease. Throughout the night, density continues to decrease, reaching a low point just before sunrise.

Four Distinct Layers

The ionosphere is composed of three layers designated D, E, and F, from lowest level to highest level as shown in figure 2-14. The F layer is further divided into two layers designated F1 (the lower layer) and F2 (the higher layer). The presence or absence of these layers in the ionosphere and their height above the Earth varies with the position of the sun. At high noon, radiation in the ionosphere directly above a given point is greatest. At night it is minimum. When the radiation is removed, many of the particles that were ionized recombine. The time interval between these conditions finds the position and number of the ionized layers within the ionosphere changing. Since the position of the sun varies daily, monthly, and yearly, with respect to a specified point on Earth, the exact position and number of layers present are extremely difficult to determine. However, the following general statements can be made:

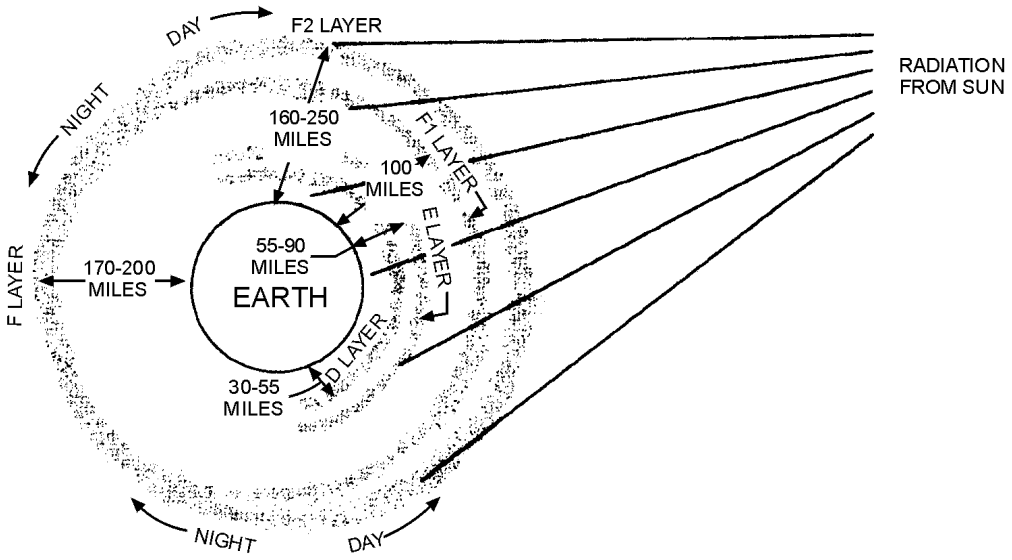


Figure 2-14.—Layers of the ionosphere.

a. The D layer ranges from about 30 to 55 miles. Ionization in the D layer is low because it is the lowest region of the ionosphere. This layer has the ability to refract signals of low frequencies. High frequencies pass right through it and are attenuated. After sunset, the D layer disappears because of the rapid recombination of ions.

b. The E layer limits are from about 55 to 90 miles. This layer is also known as the Kennelly-Heaviside layer, because these two men were the first to propose its existence. The rate of ionic recombination in this layer is rather rapid after sunset and the layer is almost gone by midnight. This layer has the ability to refract signals as high as 20 megahertz. For this reason, it is valuable for communications in ranges up to about 1500 miles.

c. The F layer exists from about 90 to 240 miles. During the daylight hours, the F layer separates into two layers, the F1 and F2 layers. The ionization level in these layers is quite high and varies widely during the day. At noon, this portion of the atmosphere is closest to the sun and the degree of ionization is maximum. Since the atmosphere is rarefied at these heights, recombination occurs slowly after sunset. Therefore, a fairly constant ionized layer is always present. The F layers are responsible for high-frequency, long distance transmission.

Q17. What causes ionization to occur in the ionosphere?

Q18. How are the four distinct layers of the ionosphere designated?

Q19. What is the height of the individual layers of the ionosphere?

REFRACTION IN THE IONOSPHERE

When a radio wave is transmitted into an ionized layer, refraction, or bending of the wave, occurs. As we discussed earlier, refraction is caused by an abrupt change in the velocity of the upper part of a radio wave as it strikes or enters a new medium. The amount of refraction that occurs depends on three main factors: (1) the density of ionization of the layer, (2) the frequency of the radio wave, and (3) the angle at which the wave enters the layer.

Density of Layer

Figure 2-15 illustrates the relationship between radio waves and ionization density. Each ionized layer has a central region of relatively dense ionization, which tapers off in intensity both above and below the maximum region. As a radio wave enters a region of INCREASING ionization, the increase in velocity of the upper part of the wave causes it to be bent back TOWARD the Earth. While the wave is in the highly dense center portion of the layer, however, refraction occurs more slowly because the density of ionization is almost uniform. As the wave enters into the upper part of the layer of DECREASING ionization, the velocity of the upper part of the wave decreases, and the wave is bent AWAY from the Earth.

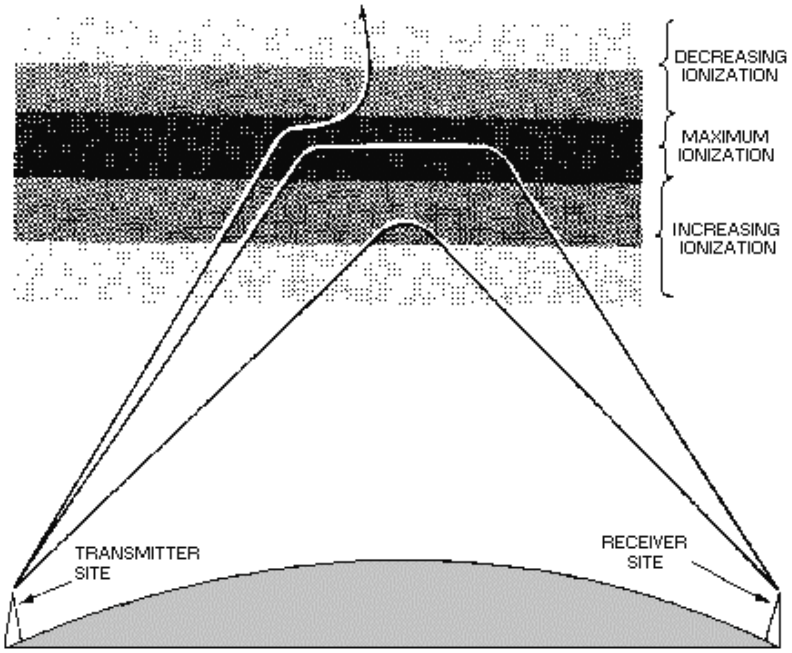


Figure 2-15.—Effects of ionospheric density on radio waves.

If a wave strikes a thin, very highly ionized layer, the wave may be bent back so rapidly that it will appear to have been reflected instead of refracted back to Earth. To reflect a radio wave, the highly ionized layer must be approximately no thicker than one wavelength of the radio wave. Since the ionized layers are often several miles thick, ionospheric reflection is more likely to occur at long wavelengths (low frequencies).

Frequency

For any given time, each ionospheric layer has a maximum frequency at which radio waves can be transmitted vertically and refracted back to Earth. This frequency is known as the CRITICAL FREQUENCY. It is a term that you will hear frequently in any discussion of radio wave propagation. Radio waves transmitted at frequencies higher than the critical frequency of a given layer will pass through the layer and be lost in space; but if these same waves enter an upper layer with a higher critical frequency, they will be refracted back to Earth. Radio waves of frequencies lower than the critical frequency will also be refracted back to Earth unless they are absorbed or have been refracted from a

lower layer. The lower the frequency of a radio wave, the more rapidly the wave is refracted by a given degree of ionization. Figure 2-16 shows three separate waves of different frequencies entering an ionospheric layer at the same angle. Notice that the 5-megahertz wave is refracted quite sharply. The 20-megahertz wave is refracted less sharply and returned to Earth at a greater distance. The 100-megahertz wave is obviously greater than the critical frequency for that ionized layer and, therefore, is not refracted but is passed into space.

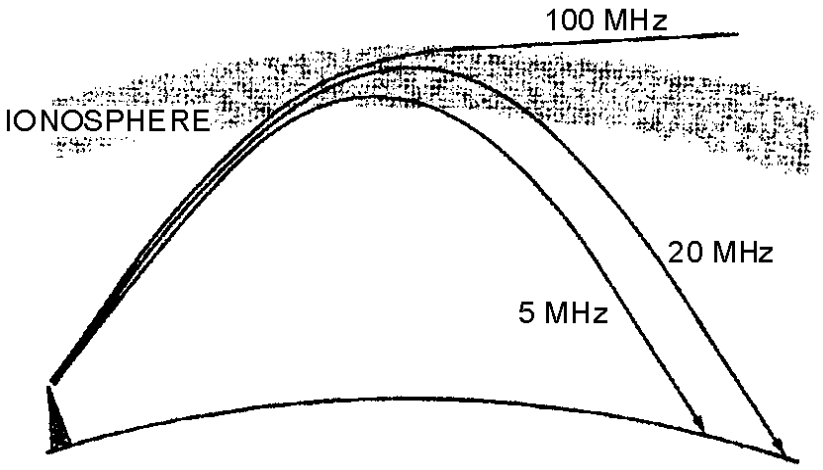


Figure 2-16.—Frequency versus refraction and distance.

Angle of Incidence

The rate at which a wave of a given frequency is refracted by an ionized layer depends on the angle at which the wave enters the layer. Figure 2-17 shows three radio waves of the same frequency entering a layer at different angles. The angle at which wave A strikes the layer is too nearly vertical for the wave to be refracted to Earth. As the wave enters the layer, it is bent slightly but passes through the layer and is lost. When the wave is reduced to an angle that is less than vertical (wave B), it strikes the layer and is refracted back to Earth. The angle made by wave B is called the CRITICAL ANGLE for that particular frequency. Any wave that leaves the antenna at an angle greater than the critical angle will penetrate the ionospheric layer for that frequency and then be lost in space. Wave C strikes the ionosphere at the smallest angle at which the wave can be refracted and still return to Earth. At any smaller angle, the wave will be refracted but will not return to Earth.

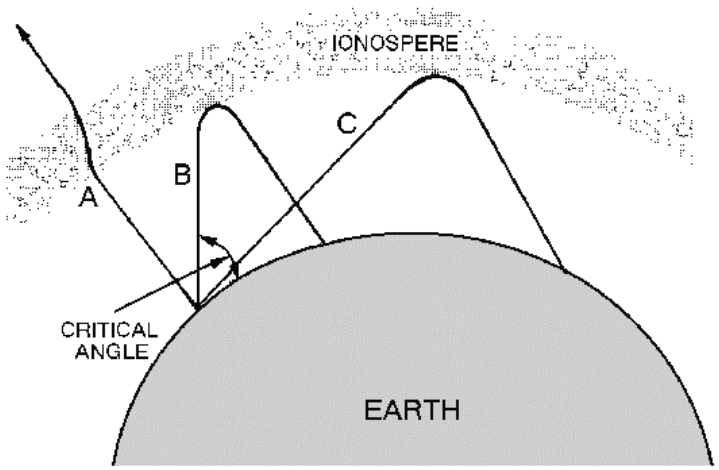


Figure 2-17.—Different incident angles of radio waves.

As the frequency of the radio wave is increased, the critical angle must be reduced for refraction to occur. This is illustrated in figure 2-18. The 2-megahertz wave strikes the layer at the critical angle for that frequency and is refracted back to Earth. Although the 5-megahertz wave (broken line) strikes the ionosphere at a lesser angle, it nevertheless penetrates the layer and is lost. As the angle is lowered from the vertical, however, a critical angle for the 5-megahertz wave is reached, and the wave is then refracted to Earth.

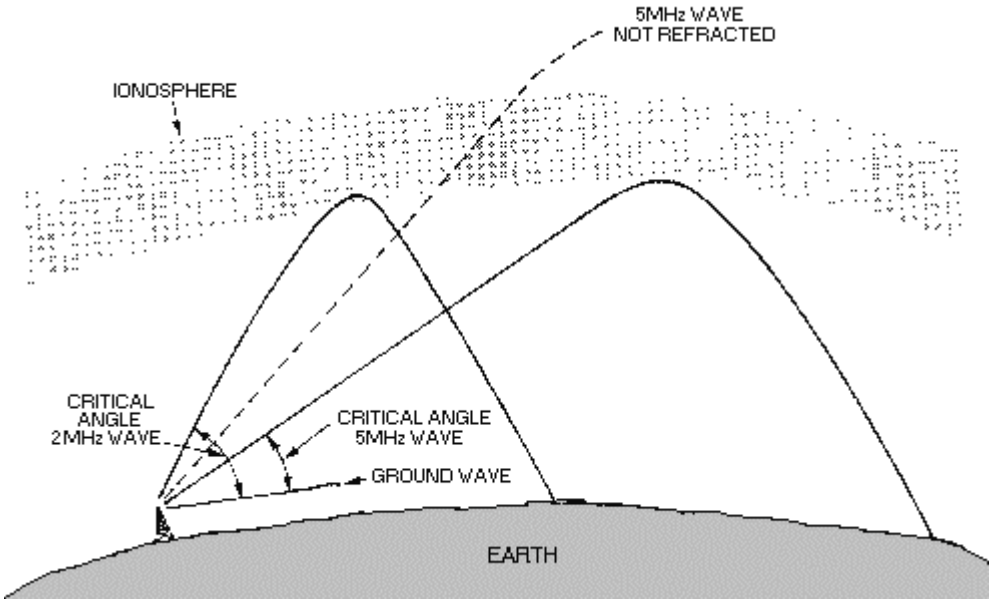


Figure 2-18.—Effects of frequency on the critical angle.

Q20. What factor determines whether a radio wave is reflected or refracted by the ionosphere?

Q21. There is a maximum frequency at which vertically transmitted radio waves can be refracted back to Earth. What is this maximum frequency called?

Q22. What three main factors determine the amount of refraction in the ionosphere?

Skip Distance/Skip Zone

In figure 2-19, note the relationship between the sky wave skip distance, the skip zone, and the ground wave coverage. The SKIP DISTANCE is the distance from the transmitter to the point where the sky wave is first returned to Earth. The size of the skip distance depends on the frequency of the wave, the angle of incidence, and the degree of ionization present.

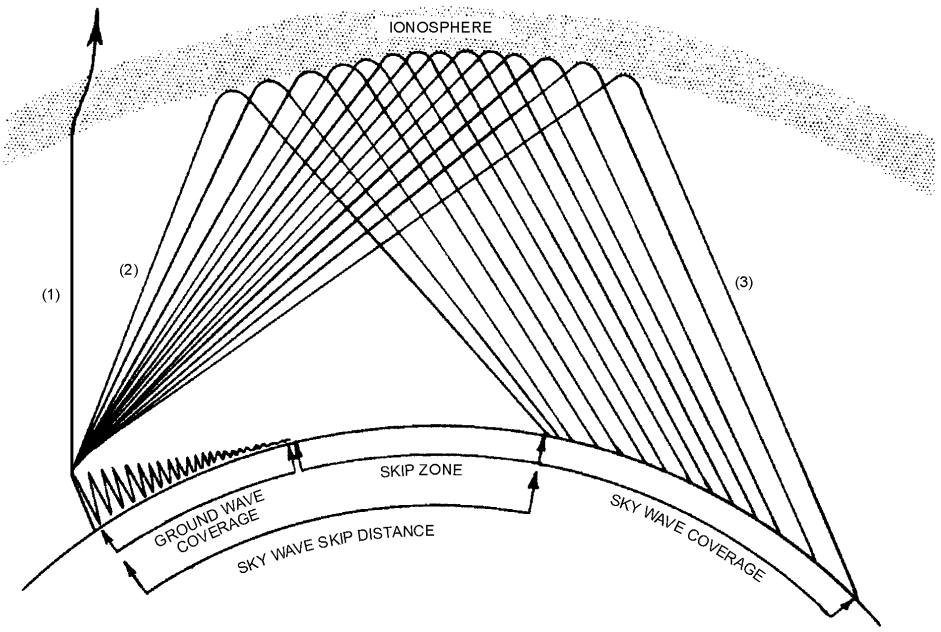


Figure 2-19.—Relationship between skip zone, skip distance, and ground wave.

The SKIP ZONE is a zone of silence between the point where the ground wave becomes too weak for reception and the point where the sky wave is first returned to Earth. The size of the skip zone depends on the extent of the ground wave coverage and the skip distance. When the ground wave coverage is great enough or the skip distance is short enough that no zone of silence occurs, there is no skip zone.

Occasionally, the first sky wave will return to Earth within the range of the ground wave. If the sky wave and ground wave are nearly of equal intensity, the sky wave alternately reinforces and cancels the ground wave, causing severe fading. This is caused by the phase difference between the two waves, a result of the longer path traveled by the sky wave.

PROPAGATION PATHS

The path that a refracted wave follows to the receiver depends on the angle at which the wave strikes the ionosphere. You should remember, however, that the rf energy radiated by a transmitting antenna spreads out with distance. The energy therefore strikes the ionosphere at many different angles rather than a single angle.

After the rf energy of a given frequency enters an ionospheric region, the paths that this energy might follow are many. It may reach the receiving antenna via two or more paths through a single layer. It

may also, reach the receiving antenna over a path involving more than one layer, by multiple hops between the ionosphere and Earth, or by any combination of these paths.

Figure 2-20 shows how radio waves may reach a receiver via several paths through one layer. The various angles at which rf energy strikes the layer are represented by dark lines and designated as rays 1 through 6.

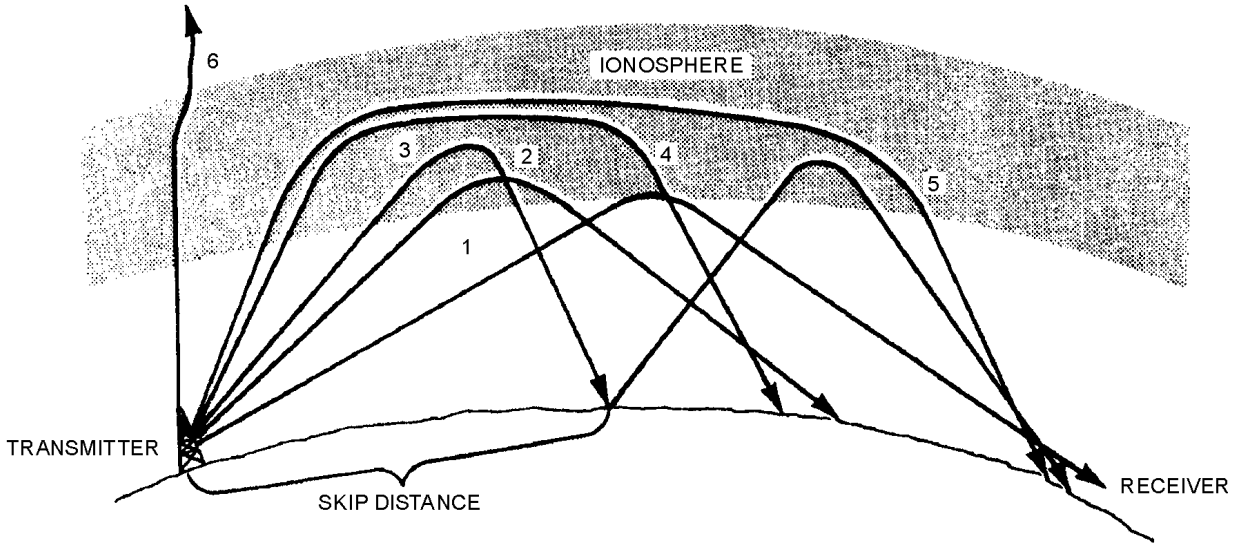


Figure 2-20.—Ray paths for a fixed frequency with varying angles of incidence.

When the angle is relatively low with respect to the horizon (ray 1), there is only slight penetration of the layer and the propagation path is long. When the angle of incidence is increased (rays 2 and 3), the rays penetrate deeper into the layer but the range of these rays decreases. When a certain angle is reached (ray 3), the penetration of the layer and rate of refraction are such that the ray is first returned to Earth at a minimal distance from the transmitter. Notice, however, that ray 3 still manages to reach the receiving site on its second refraction (called a hop) from the ionospheric layer.

As the angle is increased still more (rays 4 and 5), the rf energy penetrates the central area of maximum ionization of the layer. These rays are refracted rather slowly and are eventually returned to Earth at great distances. As the angle approaches vertical incidence (ray 6), the ray is not returned at all, but passes on through the layer.

ABSORPTION IN THE IONOSPHERE

Many factors affect a radio wave in its path between the transmitting and receiving sites. The factor that has the greatest adverse effect on radio waves is ABSORPTION. Absorption results in the loss of energy of a radio wave and has a pronounced effect on both the strength of received signals and the ability to communicate over long distances.

You learned earlier in the section on ground waves that surface waves suffer most of their absorption losses because of ground-induced voltage. Sky waves, on the other hand, suffer most of their absorption losses because of conditions in the ionosphere. Note that some absorption of sky waves may also occur at lower atmospheric levels because of the presence of water and water vapor. However, this becomes important only at frequencies above 10,000 megahertz.

Most ionospheric absorption occurs in the lower regions of the ionosphere where ionization density is greatest. As a radio wave passes into the ionosphere, it loses some of its energy to the free electrons and ions. If these high-energy free electrons and ions do not collide with gas molecules of low energy, most of the energy lost by the radio wave is reconverted into electromagnetic energy, and the wave continues to be propagated with little change in intensity. However, if the high-energy free electrons and ions do collide with other particles, much of this energy is lost, resulting in absorption of the energy from the wave. Since absorption of energy depends on collision of the particles, the greater the density of the ionized layer, the greater the probability of collisions; therefore, the greater the absorption. The highly dense D and E layers provide the greatest absorption of radio waves.

Because the amount of absorption of the sky wave depends on the density of the ionosphere, which varies with seasonal and daily conditions, it is impossible to express a fixed relationship between distance and signal strength for ionospheric propagation. Under certain conditions, the absorption of energy is so great that communicating over any distance beyond the line of sight is difficult.

FADING

The most troublesome and frustrating problem in receiving radio signals is variations in signal strength, most commonly known as FADING. There are several conditions that can produce fading. When a radio wave is refracted by the ionosphere or reflected from the Earth's surface, random changes in the polarization of the wave may occur. Vertically and horizontally mounted receiving antennas are designed to receive vertically and horizontally polarized waves, respectively. Therefore, changes in polarization cause changes in the received signal level because of the inability of the antenna to receive polarization changes.

Fading also results from absorption of the rf energy in the ionosphere. Absorption fading occurs for a longer period than other types of fading, since absorption takes place slowly.

Usually, however, fading on ionospheric circuits is mainly a result of multipath propagation.

Multipath Fading

MULTIPATH is simply a term used to describe the multiple paths a radio wave may follow between transmitter and receiver. Such propagation paths include the ground wave, ionospheric refraction, reradiation by the ionospheric layers, reflection from the Earth's surface or from more than one ionospheric layer, etc. Figure 2-21 shows a few of the paths that a signal can travel between two sites in a typical circuit. One path, XYZ, is the basic ground wave. Another path, XEA, refracts the wave at the E layer and passes it on to the receiver at A. Still another path, XFZFA, results from a greater angle of incidence and two refractions from the F layer. At point Z, the received signal is a combination of the ground wave and the sky wave. These two signals having traveled different paths arrive at point Z at different times. Thus, the arriving waves may or may not be in phase with each other. Radio waves that are received in phase reinforce each other and produce a stronger signal at the receiving site. Conversely, those that are received out of phase produce a weak or fading signal. Small alternations in the transmission path may change the phase relationship of the two signals, causing periodic fading. This condition occurs at point A. At this point, the double-hop F layer signal may be in or out of phase with the signal arriving from the E layer.

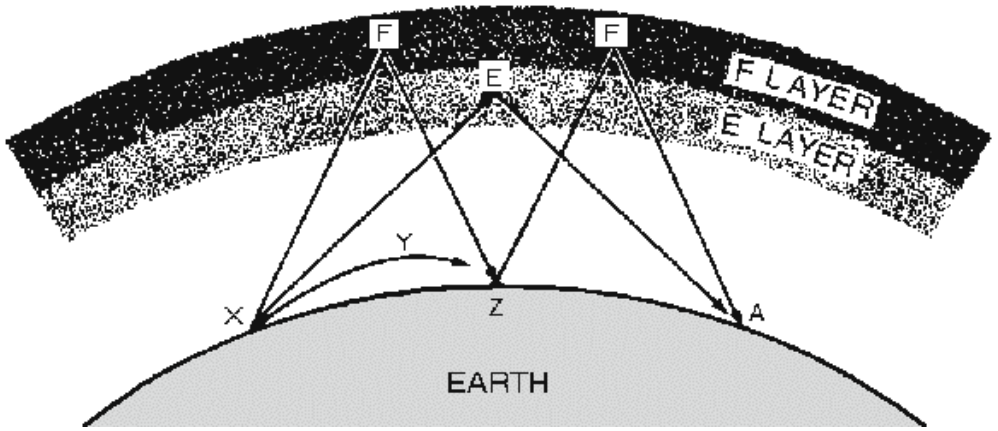


Figure 2-21.—Multipath transmission.

Multipath fading may be minimized by practices called SPACE DIVERSITY and FREQUENCY DIVERSITY. In space diversity, two or more receiving antennas are spaced some distance apart. Fading does not occur simultaneously at both antennas; therefore, enough output is almost always available from one of the antennas to provide a useful signal. In frequency diversity, two transmitters and two receivers are used, each pair tuned to a different frequency, with the same information being transmitted simultaneously over both frequencies. One of the two receivers will almost always provide a useful signal.

Selective Fading

Fading resulting from multipath propagation is variable with frequency since each frequency arrives at the receiving point via a different radio path. When a wide band of frequencies is transmitted simultaneously, each frequency will vary in the amount of fading. This variation is called SELECTIVE FADING. When selective fading occurs, all frequencies of the transmitted signal do not retain their original phases and relative amplitudes. This fading causes severe distortion of the signal and limits the total signal transmitted.

Q23. What is the skip zone of a radio wave?

Q24. Where does the greatest amount of ionospheric absorption occur in the ionosphere?

Q25. What is meant by the term "multipath"?

Q26. When a wide band of frequencies is transmitted simultaneously, each frequency will vary in the amount of fading. What is this variable fading called?

TRANSMISSION LOSSES

All radio waves propagated over ionospheric paths undergo energy losses before arriving at the receiving site. As we discussed earlier, absorption in the ionosphere and lower atmospheric levels account for a large part of these energy losses. There are two other types of losses that also significantly affect the ionospheric propagation of radio waves. These losses are known as ground reflection loss and free space loss. The combined effects of absorption, ground reflection loss, and free space loss account for most of the energy losses of radio transmissions propagated by the ionosphere.

Ground Reflection Loss

When propagation is accomplished via multihop refraction, rf energy is lost each time the radio wave is reflected from the Earth's surface. The amount of energy lost depends on the frequency of the wave, the angle of incidence, ground irregularities, and the electrical conductivity of the point of reflection.

Free space Loss

Normally, the major loss of energy is because of the spreading out of the wavefront as it travels away from the transmitter. As the distance increases, the area of the wavefront spreads out, much like the beam of a flashlight. This means the amount of energy contained within any unit of area on the wavefront will decrease as distance increases. By the time the energy arrives at the receiving antenna, the wavefront is so spread out that the receiving antenna extends into only a very small fraction of the wavefront. This is illustrated in figure 2-22.

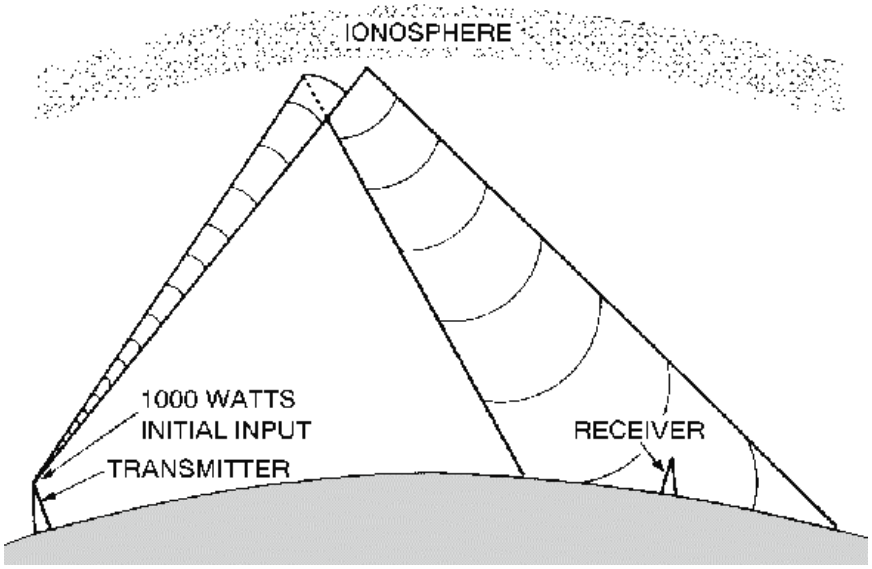


Figure 2-22.—Free space loss principle.

ELECTROMAGNETIC INTERFERENCE (EMI)

The transmission losses just discussed are not the only factors that interfere with communications. An additional factor that can interfere with radio communications is the presence of ELECTROMAGNETIC INTERFERENCE (EMI). This interference can result in annoying or impossible operating conditions. Sources of emi are both man-made and natural.

Man-Made Interference

Man-made interference may come from several sources. Some of these sources, such as oscillators, communications transmitters, and radio transmitters, may be specifically designed to generate radio frequency energy. Some electrical devices also generate radio frequency energy, although they are not specifically designed for this purpose. Examples are ignition systems, generators, motors, switches, relays, and voltage regulators. The intensity of man-made interference may vary throughout the day and drop off to a low level at night when many of these sources are not being used. Man-made interference may be a critical limiting factor at radio receiving sites located near industrial areas.

Natural Interference

Natural interference refers to the static that you often hear when listening to a radio. This interference is generated by natural phenomena, such as thunderstorms, snowstorms, cosmic sources, and the sun. The energy released by these sources is transmitted to the receiving site in roughly the same manner as radio waves. As a result, when ionospheric conditions are favorable for the long distance propagation of radio waves, they are likewise favorable for the propagation of natural interference. Natural interference is very erratic, particularly in the hf band, but generally will decrease as the operating frequency is increased and wider bandwidths are used. There is little natural interference above 30 megahertz.

Control of EMI

Electromagnetic interference can be reduced or eliminated by using various suppression techniques. The amount of emi that is produced by a radio transmitter can be controlled by cutting transmitting antennas to the correct frequency, limiting bandwidth, and using electronic filtering networks and metallic shielding.

Radiated emi during transmission can be controlled by the physical separation of the transmitting and receiving antennas, the use of directional antennas, and limiting antenna bandwidth.

Q27. What are the two main sources of emi with which radio waves must compete?

Q28. Thunderstorms, snowstorms, cosmic sources, the sun, etc., are a few examples of emi sources. What type of emi comes from these sources?

Q29. Motors, switches, voltage regulators, generators, etc., are a few examples of emi sources. What type of emi comes from these sources?

Q30. What are three ways of controlling the amount of transmitter-generated emi?

Q31. What are three ways of controlling radiated emi during transmission?

VARIATIONS IN THE IONOSPHERE

Because the existence of the ionosphere is directly related to radiations emitted from the sun, the movement of the Earth about the sun or changes in the sun's activity will result in variations in the ionosphere. These variations are of two general types: (1) those which are more or less regular and occur in cycles and, therefore, can be predicted in advance with reasonable accuracy, and (2) those which are irregular as a result of abnormal behavior of the sun and, therefore, cannot be predicted in advance. Both regular and irregular variations have important effects on radio wave propagation.

Regular Variations

The regular variations that affect the extent of ionization in the ionosphere can be divided into four main classes: daily, seasonal, 11-year, and 27-day variations.

DAILY.—Daily variations in the ionosphere are a result of the 24-hour rotation of the Earth about its axis. Daily variations of the different layers (fig. 2-14) are summarized as follows:

- The D layer reflects vlf waves; is important for long range vlf communications; refracts lf and mf waves for short range communications; absorbs hf waves; has little effect on vhf and above; and disappears at night.

- In the E layer, ionization depends on the angle of the sun. The E layer refracts hf waves during the day up to 20 megahertz to distances of about 1200 miles. Ionization is greatly reduced at night.

- Structure and density of the F region depend on the time of day and the angle of the sun. This region consists of one layer during the night and splits into two layers during daylight hours.

- Ionization density of the F1 layer depends on the angle of the sun. Its main effect is to absorb hf waves passing through to the F2 layer.

- The F2 layer is the most important layer for long distance hf communications. It is a very variable layer and its height and density change with time of day, season, and sunspot activity.

SEASONAL.—Seasonal variations are the result of the Earth revolving around the sun; the relative position of the sun moves from one hemisphere to the other with changes in seasons. Seasonal variations of the D, E, and F1 layers correspond to the highest angle of the sun; thus the ionization density of these layers is greatest during the summer. The F2 layer, however, does not follow this pattern; its ionization is greatest in winter and least in summer, the reverse of what might be expected. As a result, operating frequencies for F2 layer propagation are higher in the winter than in the summer.

ELEVEN-YEAR SUN SPOT CYCLE.—One of the most notable phenomena on the surface of the sun is the appearance and disappearance of dark, irregularly shaped areas known as SUNSPOTS. The exact nature of sunspots is not known, but scientists believe they are caused by violent eruptions on the sun and are characterized by unusually strong magnetic fields. These sunspots are responsible for variations in the ionization level of the ionosphere. Sunspots can, of course, occur unexpectedly, and the life span of individual sunspots is variable; however, a regular cycle of sunspot activity has also been observed. This cycle has both a minimum and maximum level of sunspot activity that occur approximately every 11 years.

During periods of maximum sunspot activity, the ionization density of all layers increases. Because of this, absorption in the D layer increases and the critical frequencies for the E, F1, and F2 layers are higher. At these times, higher operating frequencies must be used for long distance communications.

27-DAY SUNSPOT CYCLE.—The number of sunspots in existence at any one time is continually subject to change as some disappear and new ones emerge. As the sun rotates on its own axis, these sunspots are visible at 27-day intervals, the approximate period required for the sun to make one complete rotation.

The 27-day sunspot cycle causes variations in the ionization density of the layers on a day-to-day basis. The fluctuations in the F2 layer are greater than for any other layer. For this reason, precise predictions on a day-to-day basis of the critical frequency of the F2 layer are not possible. In calculating frequencies for long-distance communications, allowances for the fluctuations of the F2 layer must be made.

Irregular Variations

Irregular variations in ionospheric conditions also have an important effect on radio wave propagation. Because these variations are irregular and unpredictable, they can drastically affect communications capabilities without any warning.

The more common irregular variations are sporadic E, sudden ionospheric disturbances, and ionospheric storms.

SPORADIC E.—Irregular cloud-like patches of unusually high ionization, called sporadic E, often form at heights near the normal E layer. Exactly what causes this phenomenon is not known, nor can its occurrence be predicted. It is known to vary significantly with latitude, and in the northern latitudes, it appears to be closely related to the aurora borealis or northern lights.

At times the sporadic E is so thin that radio waves penetrate it easily and are returned to earth by the upper layers. At other times, it extends up to several hundred miles and is heavily ionized.

These characteristics may be either harmful or helpful to radio wave propagation. For example, sporadic E may blank out the use of higher, more favorable ionospheric layers or cause additional absorption of the radio wave at some frequencies. Also, it can cause additional multipath problems and delay the arrival times of the rays of rf energy.

On the other hand, the critical frequency of the sporadic E is very high and can be greater than double the critical frequency of the normal ionospheric layers. This condition may permit the long distance transmission of signals at unusually high frequencies. It may also permit short distance communications to locations that would normally be in the skip zone.

The sporadic E can form and disappear in a short time during either the day or night. However, it usually does not occur at the same time at all transmitting or receiving stations.

SUDDEN IONOSPHERIC DISTURBANCES.—The most startling of the ionospheric irregularities is known as a SUDDEN IONOSPHERIC DISTURBANCE (sid). These disturbances may occur without warning and may prevail for any length of time, from a few minutes to several hours. When sid occurs, long distance propagation of hf radio waves is almost totally "blanked out." The immediate effect is that radio operators listening on normal frequencies are inclined to believe their receivers have gone dead.

When sid has occurred, examination of the sun has revealed a bright solar eruption. All stations lying wholly, or in part, on the sunward side of the Earth are affected. The solar eruption produces an unusually intense burst of ultraviolet light, which is not absorbed by the F2, F1, and E layers, but instead causes a sudden abnormal increase in the ionization density of the D layer. As a result, frequencies above 1 or 2 megahertz are unable to penetrate the D layer and are usually completely absorbed by the layer.

IONOSPHERIC STORMS.—Ionospheric storms are disturbances in the Earth's magnetic field. They are associated, in a manner not fully understood, with both solar eruptions and the 27-day intervals, thus corresponding to the rotation of the sun.

Scientists believe that ionospheric storms result from particle radiation from the sun. Particles radiated from a solar eruption have a slower velocity than ultraviolet light waves produced by the eruption. This would account for the 18-hour or so time difference between a sid and an ionospheric storm. An ionospheric storm that is associated with sunspot activity may begin anytime from 2 days before an active sunspot crosses the central meridian of the sun until four days after it passes the central meridian. At times, however, active sunspots have crossed the central region of the sun without any ionospheric storms occurring. Conversely, ionospheric storms have occurred when there were no visible spots on the sun and no preceding sid. As you can see, some correlation between ionospheric storms, sid, and sunspot activity is possible, but there are no hard and fast rules. Ionospheric storms can occur suddenly without warning.

The most prominent effects of ionospheric storms are a turbulent ionosphere and very erratic sky wave propagation. Critical frequencies are lower than normal, particularly for the F2 layer. Ionospheric storms affect the higher F2 layer first, reducing its ion density. Lower layers are not appreciably affected by the storms unless the disturbance is great. The practical effect of ionospheric storms is that the range of

frequencies that can be used for communications on a given circuit is much smaller than normal, and communications are possible only at the lower working frequencies.

Q32. What are the two general types of variations in the ionosphere?

Q33. What is the main difference between these two types of variations?

Q34. What are the four main classes of regular variation which affect the extent of ionization in the ionosphere?

Q35. What are the three more common types of irregular variations in the ionosphere?

FREQUENCY SELECTION CONSIDERATIONS

Up to this point, we have covered various factors that control the propagation of radio waves through the ionosphere, such as the structure of the ionosphere, the incidence angle of radio waves, operating frequencies, etc. There is a very good reason for studying radio wave propagation. You must have a thorough knowledge of radio wave propagation to exercise good judgment when you select transmitting and receiving antennas and operating frequencies. Selection of a suitable operating frequency (within the bounds of frequency allocations and availability) is of prime importance in maintaining reliable communications.

For successful communications between any two specified locations at any given time of the day, there is a maximum frequency, a lowest frequency, and an optimum frequency that can be used.

Maximum Usable Frequency

As we discussed earlier, the higher the frequency of a radio wave, the lower the rate of refraction by an ionized layer. Therefore, for a given angle of incidence and time of day, there is a maximum frequency that can be used for communications between two given locations. This frequency is known as the MAXIMUM USABLE FREQUENCY (muf).

Waves at frequencies above the muf are normally refracted so slowly that they return to Earth beyond the desired location, or pass on through the ionosphere and are lost. You should understand, however, that use of an established muf certainly does not guarantee successful communications between a transmitting site and a receiving site. Variations in the ionosphere may occur at any time and consequently raise or lower the predetermined muf. This is particularly true for radio waves being refracted by the highly variable F2 layer.

The muf is highest around noon when ultraviolet light waves from the sun are the most intense. It then drops rather sharply as recombination begins to take place.

Lowest Usable Frequency

As there is a maximum operating frequency that can be used for communications between two points, there is also a minimum operating frequency. This is known as the LOWEST USABLE FREQUENCY (luf).

As the frequency of a radio wave is lowered, the rate of refraction increases. So a wave whose frequency is below the established luf is refracted back to Earth at a shorter distance than desired, as shown in figure 2-23.

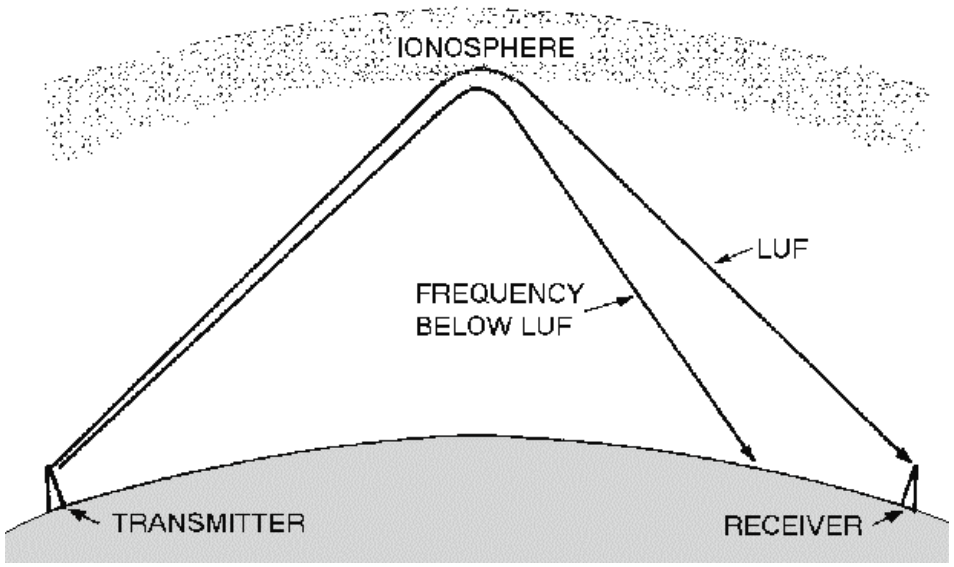


Figure 2-23.—Refraction of frequency below the lowest usable frequency (luf).

The transmission path that results from the rate of refraction is not the only factor that determines the luf. As a frequency is lowered, absorption of the radio wave increases. A wave whose frequency is too low is absorbed to such an extent that it is too weak for reception. Likewise, atmospheric noise is greater at lower frequencies; thus, a low-frequency radio wave may have an unacceptable signal-to-noise ratio.

For a given angle of incidence and set of ionospheric conditions, the luf for successful communications between two locations depends on the refraction properties of the ionosphere, absorption considerations, and the amount of atmospheric noise present.

Optimum Working Frequency

Neither the muf nor the luf is a practical operating frequency. While radio waves at the luf can be refracted back to Earth at the desired location, the signal-to-noise ratio is still much lower than at the higher frequencies, and the probability of multipath propagation is much greater. Operating at or near the muf can result in frequent signal fading and dropouts when ionospheric variations alter the length of the transmission path.

The most practical operating frequency is one that you can rely on with the least amount of problems. It should be high enough to avoid the problems of multipath, absorption, and noise encountered at the lower frequencies; but not so high as to result in the adverse effects of rapid changes in the ionosphere.

A frequency that meets the above criteria has been established and is known as the OPTIMUM WORKING FREQUENCY. It is abbreviated "fot" from the initial letters of the French words for optimum working frequency, "frequence optimum de travail." The fot is roughly about 85 percent of the muf but the actual percentage varies and may be either considerably more or less than 85 percent.

Q36. What do the letters muf, luf, and fot stand for?

Q37. When is muf at its highest and why?

Q38. What happens to the radio wave if the luf is too low?

Q39. What are some disadvantages of operating transmitters at or near the luf?

Q40. What are some disadvantages of operating a transmitter at or near the muf?

Q41. What is fot?

WEATHER VERSUS PROPAGATION

Weather is an additional factor that affects the propagation of radio waves. In this section, we will explain how and to what extent the various weather phenomena affect wave propagation.

Wind, air temperature, and water content of the atmosphere can combine in many ways. Certain combinations can cause radio signals to be heard hundreds of miles beyond the ordinary range of radio communications. Conversely, a different combination of factors can cause such attenuation of the signal that it may not be heard even over a normally satisfactory path. Unfortunately, there are no hard and fast rules on the effects of weather on radio transmissions since the weather is extremely complex and subject to frequent change. We will, therefore, limit our discussion on the effects of weather on radio waves to general terms.

PRECIPITATION ATTENUATION

Calculating the effect of weather on radio wave propagation would be comparatively simple if there were no water or water vapor in the atmosphere. However, some form of water (vapor, liquid, or solid) is always present and must be considered in all calculations. Before we begin discussing the specific effects that individual forms of precipitation (rain, snow, fog) have on radio waves, you should understand that attenuation because of precipitation is generally proportionate to the frequency and wavelength of the radio wave. For example, rain has a pronounced effect on waves at microwave frequencies. However, rain hardly affects waves with long wavelengths (hf range and below). You can assume, then, that as the wavelength becomes shorter with increases in frequency, precipitation has an increasingly important attenuation effect on radio waves. Conversely, you can assume that as the wavelength becomes longer with decreases in frequency, precipitation has little attenuation effect.

Rain

Attenuation because of raindrops is greater than attenuation because of other forms of precipitation. Attenuation may be caused by absorption, in which the raindrop, acting as a poor dielectric, absorbs power from the radio wave and dissipates the power by heat loss or by scattering (fig. 2-24). Raindrops cause greater attenuation by scattering than by absorption at frequencies above 100 megahertz. At frequencies above 6 gigahertz, attenuation by raindrop scatter is even greater.

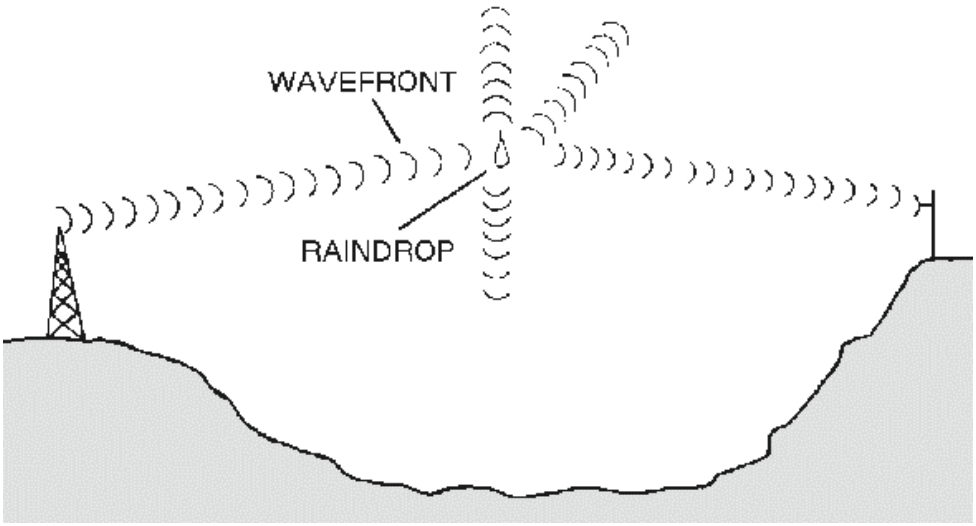


Figure 2-24.—Rf energy losses from scattering.

Fog

In the discussion of attenuation, fog may be considered as another form of rain. Since fog remains suspended in the atmosphere, the attenuation is determined by the quantity of water per unit volume and by the size of the droplets. Attenuation because of fog is of minor importance at frequencies lower than 2 gigahertz. However, fog can cause serious attenuation by absorption, at frequencies above 2 gigahertz.

Snow

The scattering effect because of snow is difficult to compute because of irregular sizes and shapes of the flakes. While information on the attenuating effect of snow is limited, scientists assume that attenuation from snow is less than from rain falling at an equal rate. This assumption is borne out by the fact that the density of rain is eight times the density of snow. As a result, rain falling at 1 inch per hour would have more water per cubic inch than snow falling at the same rate.

Hail

Attenuation by hail is determined by the size of the stones and their density. Attenuation of radio waves by scattering because of hailstones is considerably less than by rain.

TEMPERATURE INVERSION

Under normal atmospheric conditions, the warmest air is found near the surface of the Earth. The air gradually becomes cooler as altitude increases. At times, however, an unusual situation develops in which layers of warm air are formed above layers of cool air. This condition is known as TEMPERATURE INVERSION. These temperature inversions cause channels, or ducts, of cool air to be sandwiched between the surface of the Earth and a layer of warm air, or between two layers of warm air.

If a transmitting antenna extends into such a duct of cool air, or if the radio wave enters the duct at a very low angle of incidence, vhf and uhf transmissions may be propagated far beyond normal line-of-sight distances. When ducts are present as a result of temperature inversions, good reception of vhf and uhf television signals from a station located hundreds of miles away is not unusual. These long

distances are possible because of the different densities and refractive qualities of warm and cool air. The sudden change in density when a radio wave enters the warm air above a duct causes the wave to be refracted back toward Earth. When the wave strikes the Earth or a warm layer below the duct, it is again reflected or refracted upward and proceeds on through the duct with a multiple-hop type of action. An example of the propagation of radio waves by ducting is shown in figure 2-25.

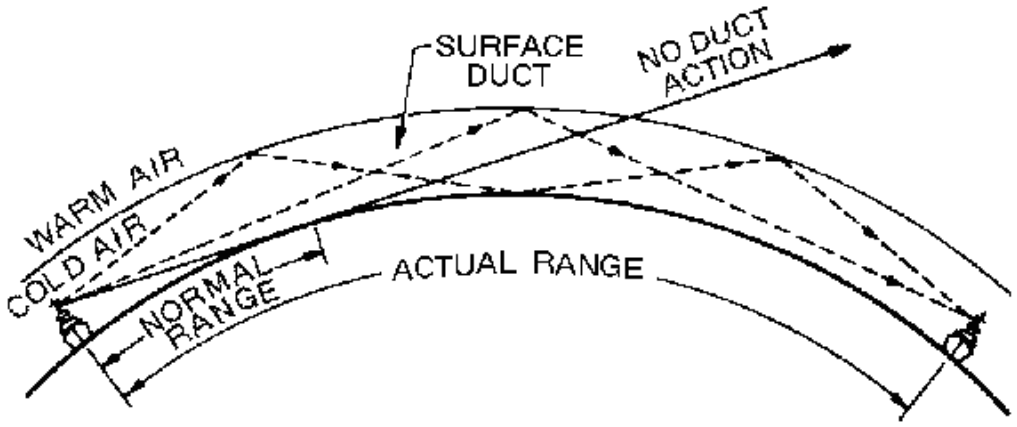


Figure 2-25.—Duct effect caused by temperature inversion.

Q42. How do raindrops affect radio waves?

Q43. How does fog affect radio waves at frequencies above 2 gigahertz?

Q44. How is the term "temperature inversion" used when referring to radio waves?

Q45. How does temperature inversion affect radio transmission?

TROPOSPHERIC PROPAGATION

As the lowest region of the Earth's atmosphere, the troposphere extends from the Earth's surface to a height of slightly over 7 miles. Virtually all weather phenomena occur in this region. Generally, the troposphere is characterized by a steady decrease in both temperature and pressure as height is increased. However, the many changes in weather phenomena cause variations in humidity and an uneven heating of the Earth's surface. As a result, the air in the troposphere is in constant motion. This motion causes small turbulences, or eddies, to be formed, as shown by the bouncing of aircraft entering turbulent areas of the atmosphere. These turbulences are most intense near the Earth's surface and gradually diminish with height. They have a refractive quality that permits the refracting or scattering of radio waves with short wavelengths. This scattering provides enhanced communications at higher frequencies.

Recall that in the relationship between frequency and wavelength, wavelength decreases as frequency increases and vice versa. Radio waves of frequencies below 30 megahertz normally have wavelengths longer than the size of weather turbulences. These radio waves are, therefore, affected very little by the turbulences. On the other hand, as the frequency increases into the vhf range and above, the wavelengths decrease in size, to the point that they become subject to tropospheric scattering. The usable frequency range for tropospheric scattering is from about 100 megahertz to 10 gigahertz.

TROPOSPHERIC SCATTERING

When a radio wave passing through the troposphere meets a turbulence, it makes an abrupt change in velocity. This causes a small amount of the energy to be scattered in a forward direction and returned to Earth at distances beyond the horizon. This phenomenon is repeated as the radio wave meets other turbulences in its path. The total received signal is an accumulation of the energy received from each of the turbulences.

This scattering mode of propagation enables vhf and uhf signals to be transmitted far beyond the normal line-of-sight. To better understand how these signals are transmitted over greater distances, you must first consider the propagation characteristics of the space wave used in vhf and uhf line-of-sight communications. When the space wave is transmitted, it undergoes very little attenuation within the line-of-sight horizon. When it reaches the horizon, the wave is diffracted and follows the Earth's curvature. Beyond the horizon, the rate of attenuation increases very rapidly and signals soon become very weak and unusable.

Tropospheric scattering, on the other hand, provides a usable signal at distances beyond the point where the diffracted space wave drops to an unusable level. This is because of the height at which scattering takes place. The turbulence that causes the scattering can be visualized as a relay station located above the horizon; it receives the transmitted energy and then reradiates it in a forward direction to some point beyond the line-of-sight distance. A high gain receiving antenna aimed toward this scattered energy can then capture it.

The magnitude of the received signal depends on the number of turbulences causing scatter in the desired direction and the gain of the receiving antenna. The scatter area used for tropospheric scatter is known as the *scatter volume*. The angle at which the receiving antenna must be aimed to capture the scattered energy is called the *scatter angle*. The scatter volume and scatter angle are shown in figure 2-26.

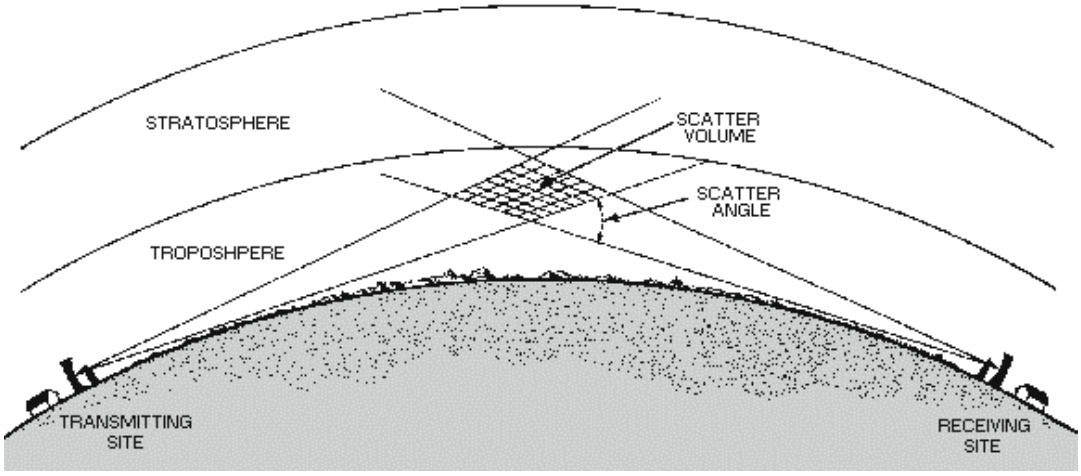


Figure 2-26.—Tropospheric scattering propagation.

The signal take-off angle (transmitting antenna's angle of radiation) determines the height of the scatter volume and the size of the scatter angle. A low signal take-off angle produces a low scatter volume, which in turn permits a receiving antenna that is aimed at a low angle to the scatter volume to capture the scattered energy.

As the signal take-off angle is increased, the height of the scatter volume is increased. When this occurs, the amount of received energy decreases. There are two reasons for this: (1) scatter angle

increases as the height of the scatter volume is increased; (2) the amount of turbulence decreases with height. As the distance between the transmitting and receiving antennas is increased, the height of the scatter volume must also be increased. The received signal level, therefore, decreases as circuit distance is increased.

The tropospheric region that contributes most strongly to tropospheric scatter propagation lies near the midpoint between the transmitting and receiving antennas and just above the radio horizon of the antennas.

Since tropospheric scatter depends on turbulence in the atmosphere, changes in atmospheric conditions have an effect on the strength of the received signal. Both daily and seasonal variations in signal strength occur as a result of changes in the atmosphere. These variations are called *long-term fading*.

In addition to long-term fading, the tropospheric scatter signal often is characterized by very rapid fading because of multipath propagation. Since the turbulent condition is constantly changing, the path lengths and individual signal levels are also changing, resulting in a rapidly changing signal. Although the signal level of the received signal is constantly changing, the average signal level is stable; therefore, no complete fade out occurs.

Another characteristic of a tropospheric scatter signal is its relatively low power level. Since very little of the scattered energy is reradiated toward the receiver, the efficiency is very low and the signal level at the final receiver point is low. Initial input power must be high to compensate for the low efficiency in the scatter volume. This is accomplished by using high-power transmitters and high-gain antennas, which concentrate the transmitted power into a beam, thus increasing the intensity of energy of each turbulence in the volume. The receiver must also be very sensitive to detect the low-level signals.

APPLICATION OF TROPOSPHERIC SCATTERING

Tropospheric scatter propagation is used for point-to-point communications. A correctly designed tropospheric scatter circuit will provide highly reliable service for distances ranging from 50 miles to 500 miles. Tropospheric scatter systems may be particularly useful for communications to locations in rugged terrain that are difficult to reach with other methods of propagation. One reason for this is that the tropospheric scatter circuit is not affected by ionospheric and auroral disturbances.

Q46. In what layer of the atmosphere does virtually all weather phenomena occur?

Q47. Which radio frequency bands use the tropospheric scattering principle for propagation of radio waves?

Q48. Where is the tropospheric region that contributes most strongly to tropospheric scatter propagation?

SUMMARY

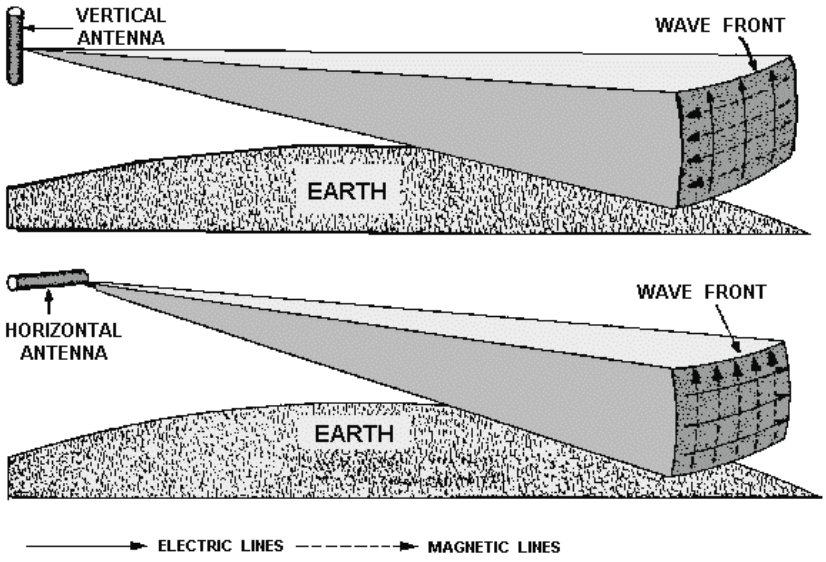
Now that you have completed this chapter, let's review some of the new terms, concepts, and ideas that you have learned. You should have a thorough understanding of these principles before moving on to chapter 3.

The **INDUCTION FIELD** contains an E field and an H field and is localized near the antenna. The E and H fields of the induction field are 90 degrees out of phase with each other.

The **RADIATION FIELD** contains E and H fields that are propagated from the antenna into space in the form of electromagnetic waves. The E and H fields of the radiation field are in phase with each other.

A **HARMONIC FREQUENCY** is any frequency that is a whole number multiple of a smaller basic frequency. For example, a radio wave transmitted at a fundamental frequency of 3000 hertz can have a second harmonic of 6000 hertz, a third harmonic frequency of 9000 hertz, etc., transmitted at the same time.

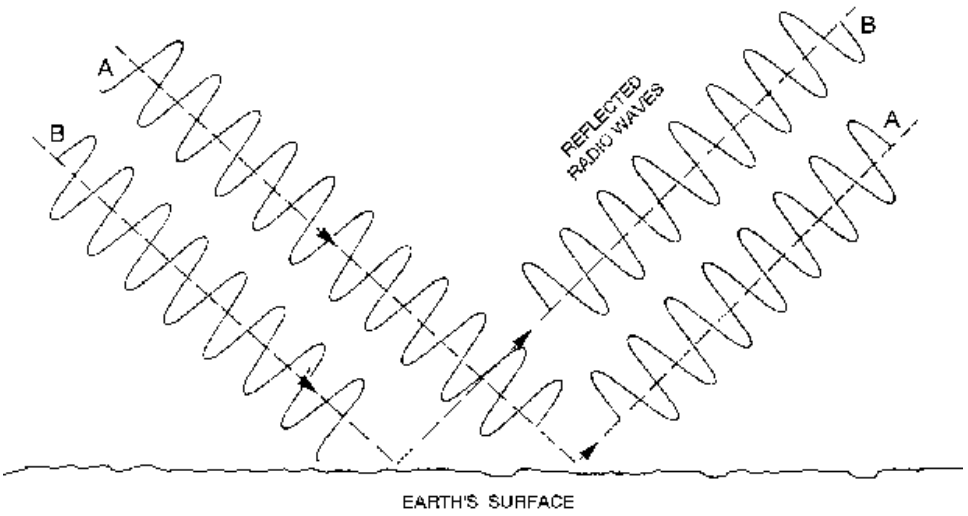
A **VERTICALLY POLARIZED** antenna transmits an electromagnetic wave with the E field perpendicular to the Earth's surface. A **HORIZONTALLY POLARIZED** antenna transmits a radio wave with the E field parallel to the Earth's surface.



A **WAVEFRONT** is a small section of an expanding sphere of radiated energy and is perpendicular to the direction of travel from the antenna.

RADIO WAVES are electromagnetic waves that can be reflected, refracted, and diffracted in the atmosphere like light and heat waves.

REFLECTED RADIO WAVES are waves that have been reflected from a surface and are 180 degrees out of phase with the initial wave.



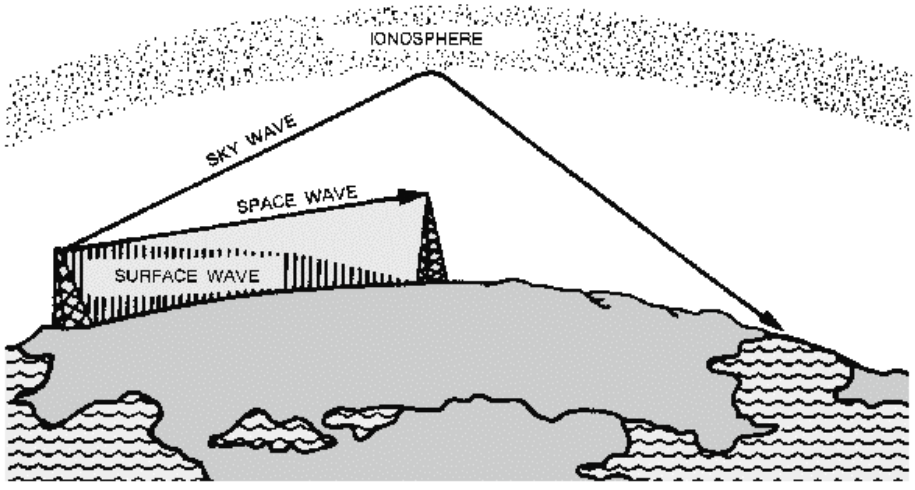
The Earth's atmosphere is divided into three separate layers: The **TROPOSPHERE**, **STRATOSPHERE**, and **IONOSPHERE**.

The **TROPOSPHERE** is the region of the atmosphere where virtually all weather phenomena take place. In this region, rf energy is greatly affected.

The **STRATOSPHERE** has a constant temperature and has little effect on radio waves.

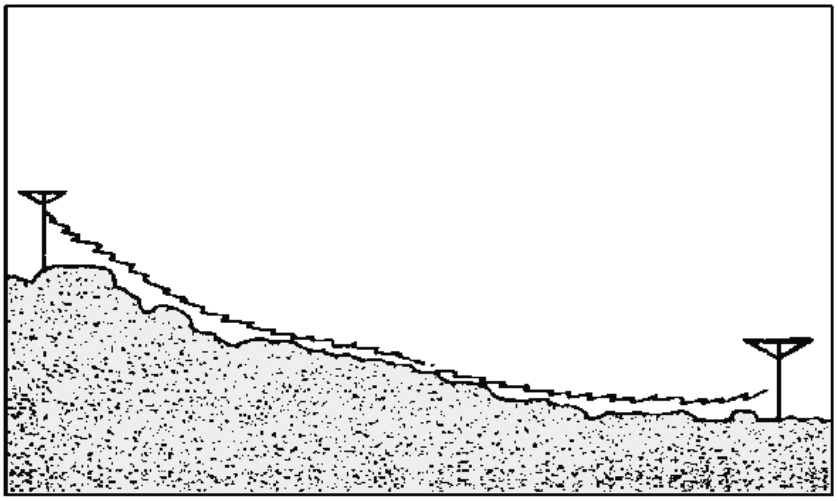
The **IONOSPHERE** contains four cloud-like layers of electrically charged ions which aid in long distance communications.

GROUND WAVES and **SKY WAVES** are the two basic types of radio waves that transmit energy from the transmitting antenna to the receiving antenna.

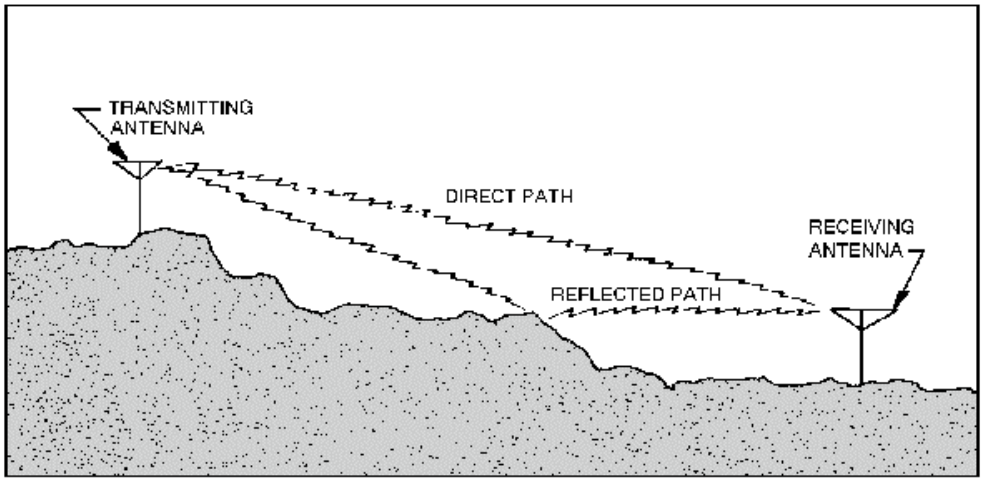


GROUND WAVES are composed of two separate component waves: the **SURFACE WAVE** and the **SPACE WAVE**.

SURFACE WAVES travel along the contour of the Earth by diffraction.



SPACE WAVES can travel through the air directly to the receiving antenna or can be reflected from the surface of the Earth.



SKY WAVES, often called ionospheric waves, are radiated in an upward direction and returned to Earth at some distant location because of refraction.

NATURAL HORIZON is the line-of-sight horizon.

RADIO HORIZON is one-third farther than the natural horizon.

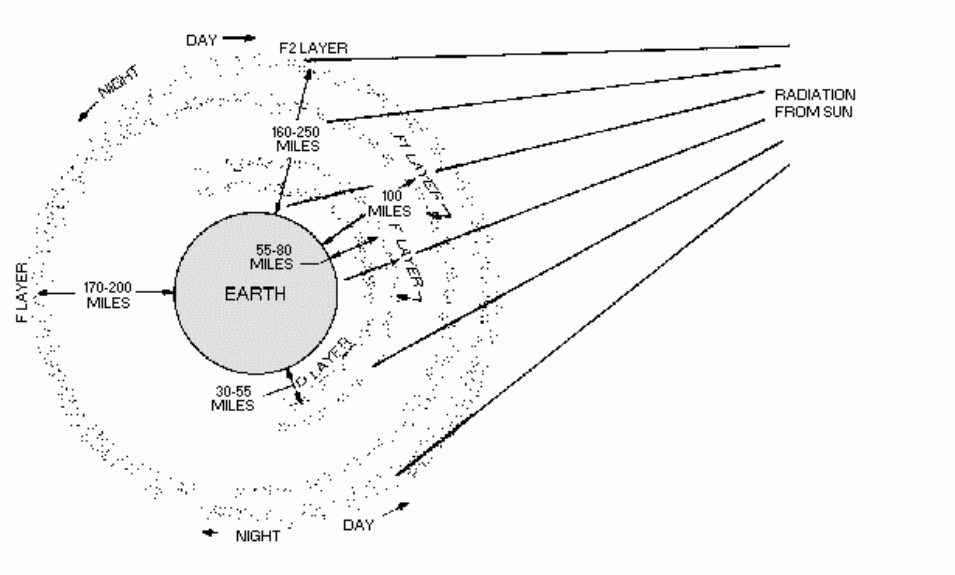
The **IONOSPHERE** consists of several layers of ions, formed by the process called ionization.

IONIZATION is the process of knocking electrons free from their parent atom, thus upsetting electrical neutrality.

RECOMBINATION is the opposite of ionization; that is, the free ions combine with positive ions, causing the positive ions to return to their original neutral atom state.

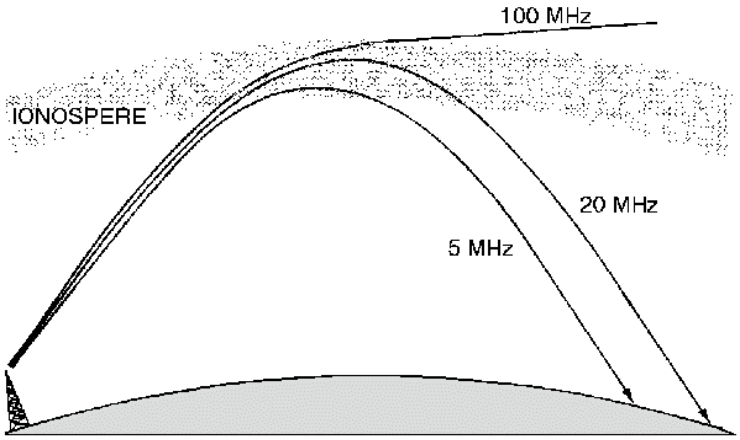
The **D LAYER** is the lowest region of the ionosphere and refracts signals of low frequencies back to Earth.

The **E LAYER** is present during the daylight hours; refracts signals as high as 20 megahertz back to Earth; and is used for communications up to 1500 miles.

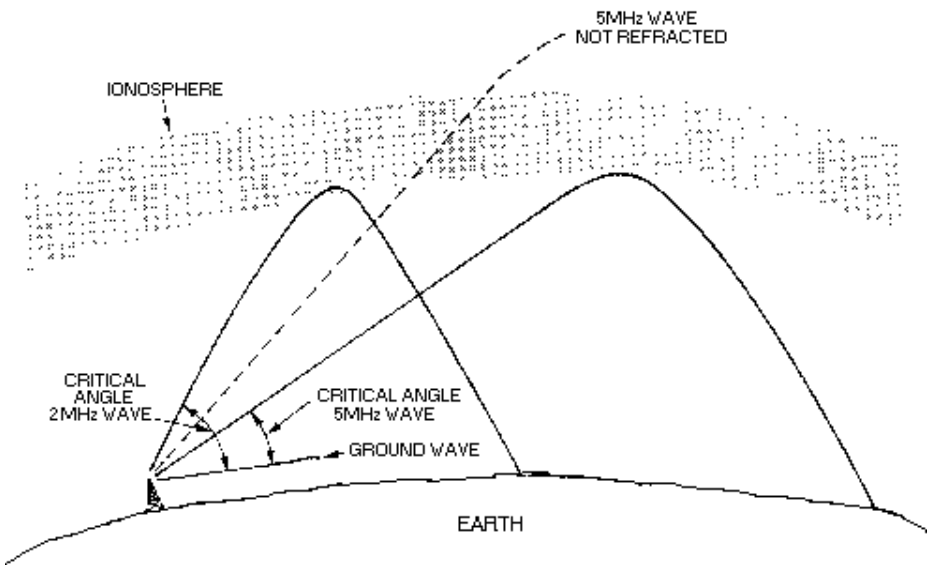


The **F LAYER** is divided into the F1 and F2 layers during the day but combine at night to form one layer. This layer is responsible for high-frequency, long-range transmission.

The **CRITICAL FREQUENCY** is the maximum frequency that a radio wave can be transmitted vertically and still be refracted back to Earth.

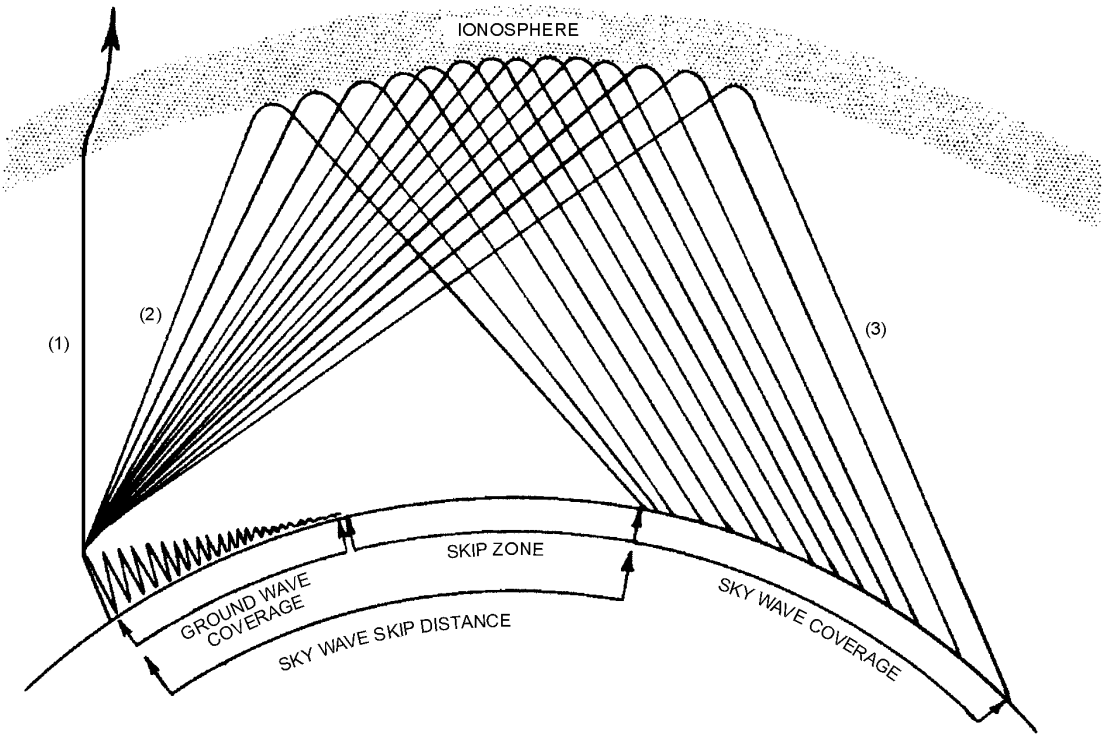


The **CRITICAL ANGLE** is the maximum and/or minimum angle that a radio wave can be transmitted and still be refracted back to Earth.



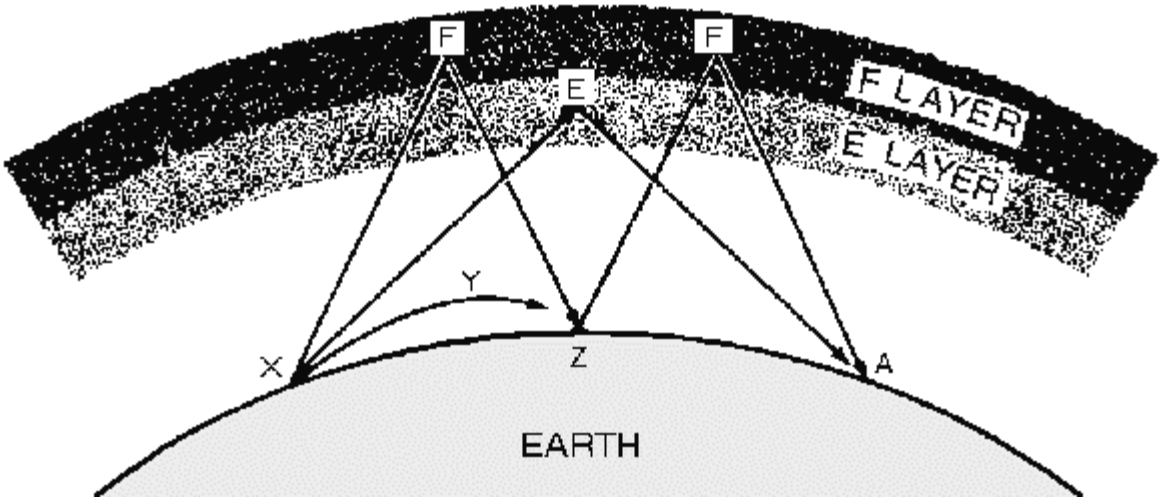
SKIP DISTANCE is the distance between the transmitter and the point where the sky wave first returns to Earth.

SKIP ZONE is the zone of silence between the point where the ground wave becomes too weak for reception and the point where the sky wave is first returned to Earth.



FADING is caused by variations in signal strength, such as absorption of the rf energy by the ionosphere.

MULTIPATH FADING occurs when a transmitted signal divides and takes more than one path to a receiver and some of the signals arrive out of phase, resulting in a weak or fading signal.

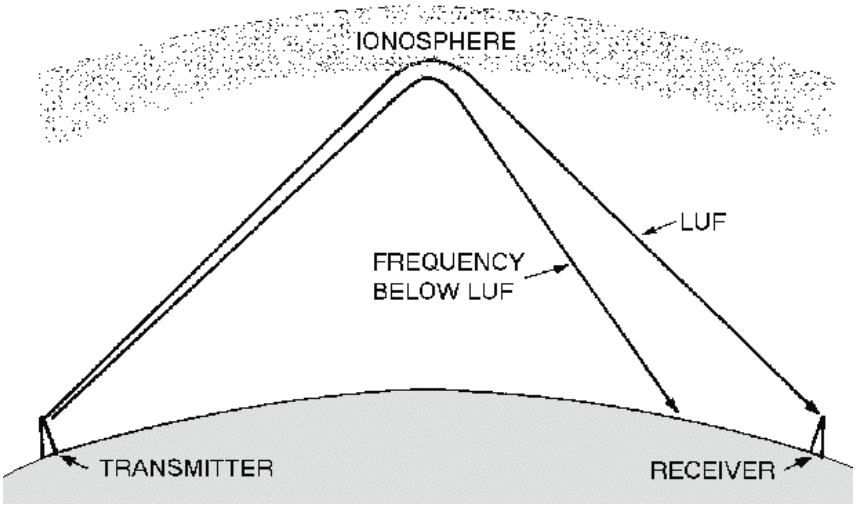


Some **TRANSMISSION LOSSES** that affect radio-wave propagation are ionospheric absorption, ground reflection, and free-space losses.

ELECTROMAGNETIC INTERFERENCE (emi), both natural and man-made, interfere with radio communications.

The **MAXIMUM USABLE FREQUENCY (muf)** is the highest frequency that can be used for communications between two locations at a given angle of incidence and time of day.

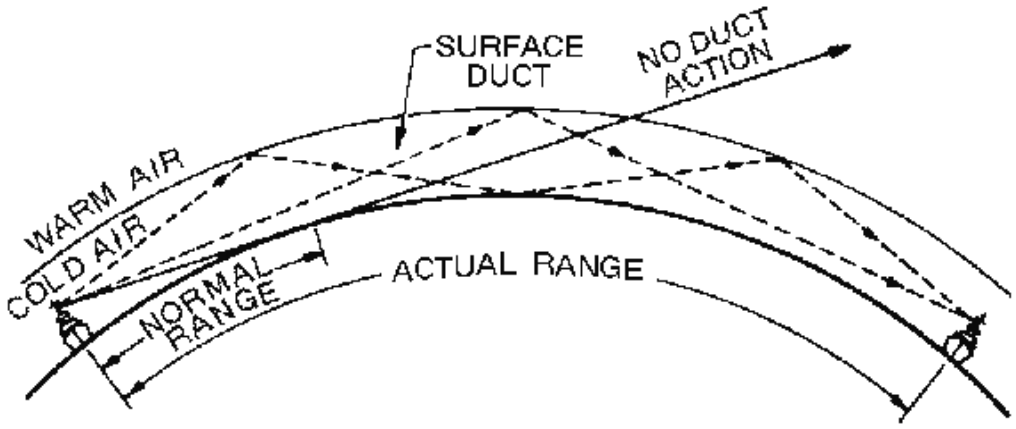
The **LOWEST USABLE FREQUENCY (luf)** is the lowest frequency that can be used for communications between two locations.



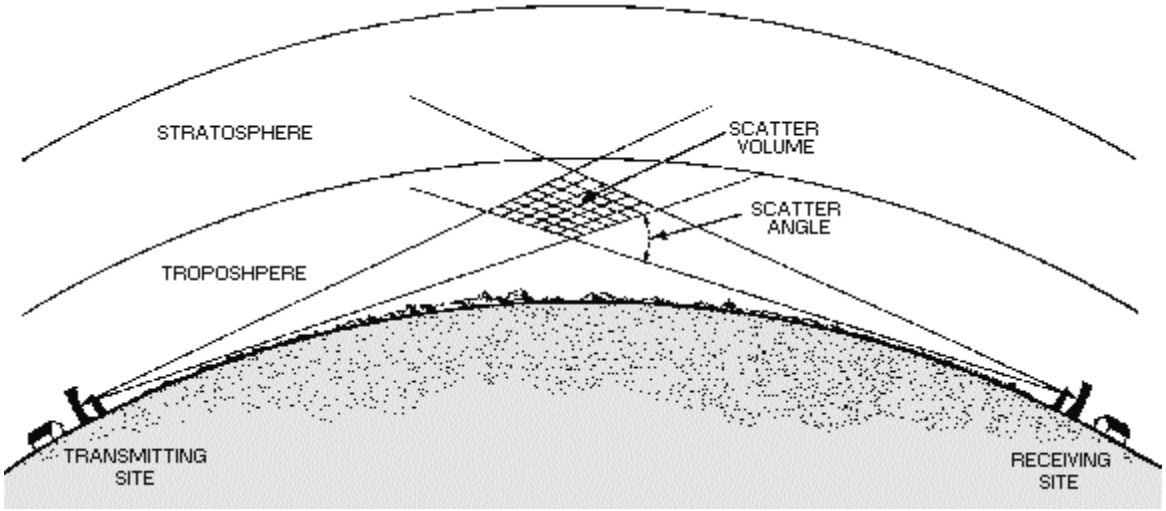
OPTIMUM WORKING FREQUENCY (fot) is the most practical operating frequency and the one that can be relied on to have the fewest problems.

PRECIPITATION ATTENUATION can be caused by rain, fog, snow, and hail; and can affect overall communications considerably.

TEMPERATURE INVERSION causes channels, or ducts, of cool air to form between layers of warm air, which can cause radio waves to travel far beyond the normal line-of-sight distances.



TROPOSPHERIC PROPAGATION uses the scattering principle to achieve beyond the line-of-sight radio communications within the troposphere.



ANSWERS TO QUESTIONS Q1. THROUGH Q48.

A1. Induction field and radiation field.

A2. Induction field.

A3. Radiation field.

A4. Fundamental frequency.

A5. Harmonic frequency or harmonics.

A6. 30 meters.

A7. 5 megahertz.

A8. Vertically polarized.

A9. Direction of wave propagation.

A10. Shifting in the phase relationships of the wave.

A11. Troposphere, stratosphere, and ionosphere.

A12. Stratosphere.

A13. Whether the component of the wave is travelling along the surface or over the surface of the earth.

A14. Radio horizon is about 1/3 farther.

A15. Sea water.

A16. (a) electrical properties of the terrain (b) frequency (c) polarization of the antenna

A17. High energy ultraviolet light waves from the sun.

A18. D, E, F_1, and F_2 layers.

A19. D layer is 30-55 miles, E layer 55-90 miles, and F layers are 90-240 miles.

A20. Thickness of ionized layer.

A21. Critical frequency.

A22. (a) density of ionization of the layer (b) frequency (c) angle at which it enters the layer

A23. A zone of silence between the ground wave and sky wave where there is no reception.

A24. Where ionization density is greatest.

A25. A term used to describe the multiple pattern a radio wave may follow.

A26. Selective fading.

A27. Natural and man-made interference.

A28. Natural.

A29. Man-made.

A30. (a) filtering and shielding of the transmitter (b) limiting bandwidth (c) cutting the antenna to the correct frequency

A31. (a) physical separation of the antenna (b) limiting bandwidth of the antenna (c) use of directional antennas

A32. Regular and irregular variations.

A33. Regular variations can be predicted but irregular variations are unpredictable.

A34. Daily, seasonal, 11-year, and 27-days variation.

A35. Sporadic E, sudden disturbances, and ionospheric storms.

A36. Muf is maximum usable frequency. Luf is lowest usable frequency. Fot is commonly known as optimum working frequency.

A37. Muf is highest around noon. Ultraviolet light waves from the sun are most intense.

A38. When luf is too low it is absorbed and is too weak for reception.

A39. Signal-to-noise ratio is low and the probability of multipath propagation is greater.

A40. Frequent signal fading and dropouts.

A41. Fot is the most practical operating frequency that can be relied on to avoid problems of multipath, absorbtion, and noise.

A42. They can cause attenuation by scattering.

A43. It can cause attenuation by absorbtion.

A44. It is a condition where layers of warm air are formed above layers of cool air.

A45. It can cause vhf and uhf transmission to be propagated far beyond normal line-of-sight distances.

A46. Troposphere.

A47. Vhf and above.

A48. Near the mid-point between the transmitting and receiving antennas, just above the radio horizon.

CHAPTER 3

PRINCIPLES OF TRANSMISSION LINES

LEARNING OBJECTIVES

Upon completion of this chapter, you will be able to:

1. State what a transmission line is and how transmission lines are used.
2. Explain the operating principles of transmission lines.
3. Describe the five types of transmission lines.
4. State the length of a transmission line.
5. Explain the theory of the transmission line.
6. Define the term LUMPED CONSTANTS in relation to a transmission line.
7. Define the term DISTRIBUTED CONSTANTS in relation to a transmission line.
8. Define LEAKAGE CURRENT.
9. Describe how the electromagnetic lines of force around a transmission line are affected by the distributed constants.
10. Define the term CHARACTERISTIC IMPEDANCE and explain how it affects the transfer of energy along a transmission line.
11. State how the energy transfer along a transmission line is affected by characteristic impedance and the infinite line.
12. Identify the cause of and describe the characteristics of reflections on a transmission line.
13. Define the term STANDING WAVES as applied to a transmission line.
14. Describe how standing waves are produced on a transmission line and identify the types of terminations.
15. Describe the types of standing-wave ratios.

INTRODUCTION TO TRANSMISSION LINES

A TRANSMISSION LINE is a device designed to guide electrical energy from one point to another. It is used, for example, to transfer the output rf energy of a transmitter to an antenna. This energy will not travel through normal electrical wire without great losses. Although the antenna can be connected directly to the transmitter, the antenna is usually located some distance away from the transmitter. On board ship,

the transmitter is located inside a radio room and its associated antenna is mounted on a mast. A transmission line is used to connect the transmitter and the antenna.

The transmission line has a single purpose for both the transmitter and the antenna. This purpose is to transfer the energy output of the transmitter to the antenna with the least possible power loss. How well this is done depends on the special physical and electrical characteristics (impedance and resistance) of the transmission line.

TERMINOLOGY

All transmission lines have two ends (see figure 3-1). The end of a two-wire transmission line connected to a source is ordinarily called the INPUT END or the GENERATOR END. Other names given to this end are TRANSMITTER END, SENDING END, and SOURCE. The other end of the line is called the OUTPUT END or RECEIVING END. Other names given to the output end are LOAD END and SINK.

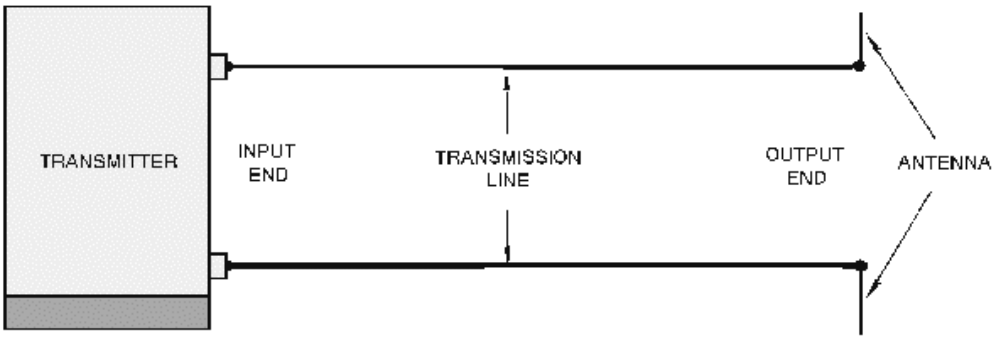


Figure 3-1.—Basic transmission line.

You can describe a transmission line in terms of its impedance. The ratio of voltage to current (E_{in}/I_{in}) at the input end is known as the INPUT IMPEDANCE (Z_{in}). This is the impedance presented to the transmitter by the transmission line and its load, the antenna. The ratio of voltage to current at the output (E_{out}/I_{out}) end is known as the OUTPUT IMPEDANCE (Z_{out}). This is the impedance presented to the load by the transmission line and its source. If an infinitely long transmission line could be used, the ratio of voltage to current at any point on that transmission line would be some particular value of impedance. This impedance is known as the CHARACTERISTIC IMPEDANCE.

Q1. What connecting link is used to transfer energy from a radio transmitter to its antenna located on the mast of a ship?

Q2. What term is used for the end of the transmission line that is connected to a transmitter?

Q3. What term is used for the end of the transmission line that is connected to an antenna?

TYPES OF TRANSMISSION MEDIUMS

The Navy uses many different types of TRANSMISSION MEDIUMS in its electronic applications. Each medium (line or wave guide) has a certain characteristic impedance value, current-carrying capacity, and physical shape and is designed to meet a particular requirement.

The five types of transmission mediums that we will discuss in this chapter include PARALLEL-LINE, TWISTED PAIR, SHIELDED PAIR, COAXIAL LINE, and WAVEGUIDES. The use of a particular line depends, among other things, on the applied frequency, the power-handling capabilities, and the type of installation.

NOTE: In the following paragraphs, we will mention LOSSES several times. We will discuss these losses more thoroughly under "LOSSES IN TRANSMISSION LINES."

Two-Wire Open Line

One type of parallel line is the TWO-WIRE OPEN LINE illustrated in figure 3-2. This line consists of two wires that are generally spaced from 2 to 6 inches apart by insulating spacers. This type of line is most often used for power lines, rural telephone lines, and telegraph lines. It is sometimes used as a transmission line between a transmitter and an antenna or between an antenna and a receiver. An advantage of this type of line is its simple construction. The principal disadvantages of this type of line are the high radiation losses and electrical noise pickup because of the lack of shielding. Radiation losses are produced by the changing fields created by the changing current in each conductor.

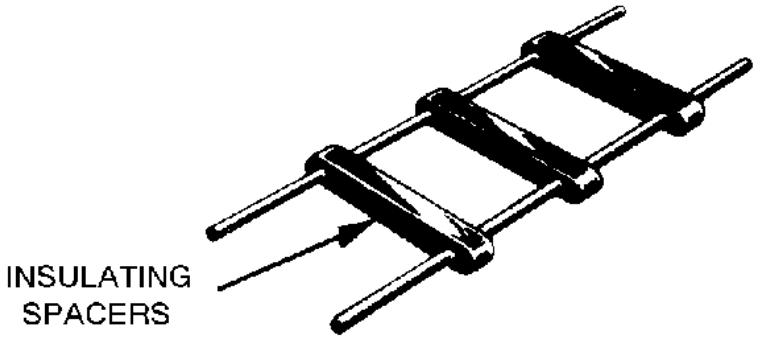


Figure 3-2.—Parallel two-wire line.

Another type of parallel line is the TWO-WIRE RIBBON (TWIN LEAD) illustrated in figure 3-3. This type of transmission line is commonly used to connect a television receiving antenna to a home television set. This line is essentially the same as the two-wire open line except that uniform spacing is assured by embedding the two wires in a low-loss dielectric, usually polyethylene. Since the wires are embedded in the thin ribbon of polyethylene, the dielectric space is partly air and partly polyethylene.

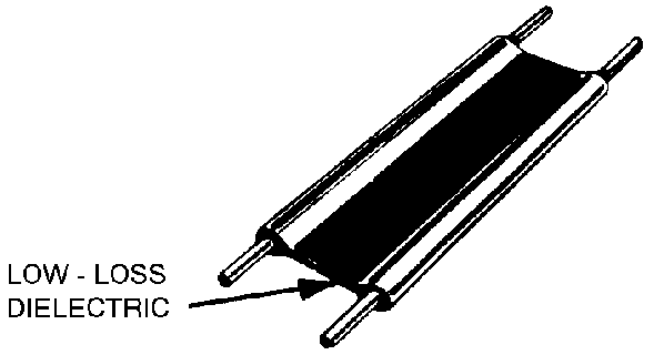


Figure 3-3.—Two-wire ribbon type line.

Twisted Pair

The TWISTED PAIR transmission line is illustrated in figure 3-4. As the name implies, the line consists of two insulated wires twisted together to form a flexible line without the use of spacers. It is not used for transmitting high frequency because of the high dielectric losses that occur in the rubber insulation. When the line is wet, the losses increase greatly.

Figure 3-4.—Twisted pair.

Shielded Pair

The SHIELDED PAIR, shown in figure 3-5, consists of parallel conductors separated from each other and surrounded by a solid dielectric. The conductors are contained within a braided copper tubing that acts as an electrical shield. The assembly is covered with a rubber or flexible composition coating that protects the line from moisture and mechanical damage. Outwardly, it looks much like the power cord of a washing machine or refrigerator.

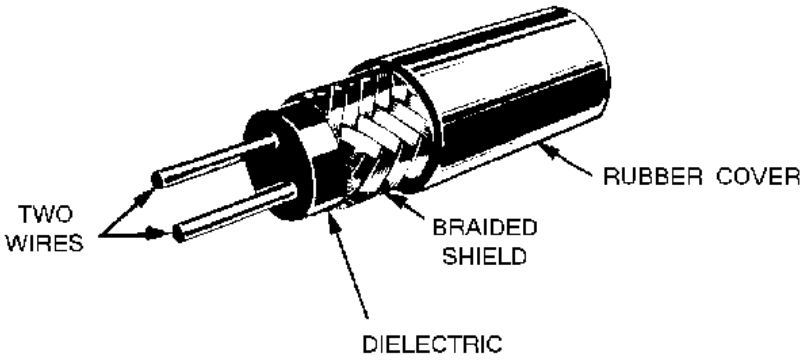


Figure 3-5.—Shielded pair.

The principal advantage of the shielded pair is that the conductors are balanced to ground; that is, the capacitance between the wires is uniform throughout the length of the line. This balance is due to the uniform spacing of the grounded shield that surrounds the wires along their entire length. The braided copper shield isolates the conductors from stray magnetic fields.

Coaxial Lines

There are two types of COAXIAL LINES, RIGID (AIR) COAXIAL LINE and FLEXIBLE (SOLID) COAXIAL LINE. The physical construction of both types is basically the same; that is, each contains two concentric conductors.

The rigid coaxial line consists of a central, insulated wire (inner conductor) mounted inside a tubular outer conductor. This line is shown in figure 3-6. In some applications, the inner conductor is also tubular. The inner conductor is insulated from the outer conductor by insulating spacers or beads at regular intervals. The spacers are made of Pyrex, polystyrene, or some other material that has good insulating characteristics and low dielectric losses at high frequencies.

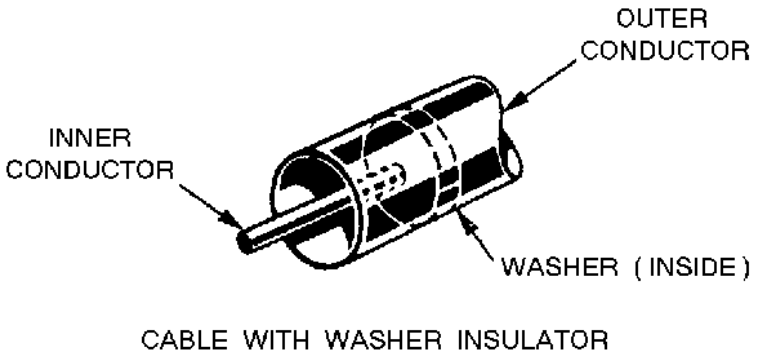


Figure 3-6.—Air coaxial line.

The chief advantage of the rigid line is its ability to minimize radiation losses. The electric and magnetic fields in a two-wire parallel line extend into space for relatively great distances and radiation losses occur. However, in a coaxial line no electric or magnetic fields extend outside of the outer conductor. The fields are confined to the space between the two conductors, resulting in a perfectly shielded coaxial line. Another advantage is that interference from other lines is reduced.

The rigid line has the following disadvantages: (1) it is expensive to construct; (2) it must be kept dry to prevent excessive leakage between the two conductors; and (3) although high-frequency losses are somewhat less than in previously mentioned lines, they are still excessive enough to limit the practical length of the line.

Leakage caused by the condensation of moisture is prevented in some rigid line applications by the use of an inert gas, such as nitrogen, helium, or argon. It is pumped into the dielectric space of the line at a pressure that can vary from 3 to 35 pounds per square inch. The inert gas is used to dry the line when it is first installed and pressure is maintained to ensure that no moisture enters the line.

Flexible coaxial lines (figure 3-7) are made with an inner conductor that consists of flexible wire insulated from the outer conductor by a solid, continuous insulating material. The outer conductor is made of metal braid, which gives the line flexibility. Early attempts at gaining flexibility involved using rubber insulators between the two conductors. However, the rubber insulators caused excessive losses at high frequencies.

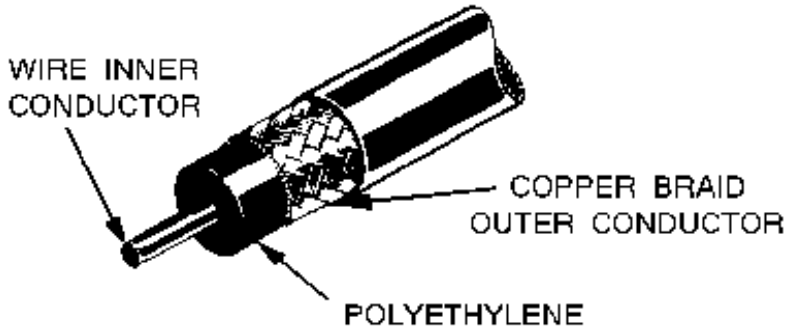


Figure 3-7.—Flexible coaxial line.

Because of the high-frequency losses associated with rubber insulators, polyethylene plastic was developed to replace rubber and eliminate these losses. Polyethylene plastic is a solid substance that remains flexible over a wide range of temperatures. It is unaffected by seawater, gasoline, oil, and most other liquids that may be found aboard ship. The use of polyethylene as an insulator results in greater high-frequency losses than the use of air as an insulator. However, these losses are still lower than the losses associated with most other solid dielectric materials.

Waveguides

The WAVEGUIDE is classified as a transmission line. However, the method by which it transmits energy down its length differs from the conventional methods. Waveguides are cylindrical, elliptical, or rectangular (cylindrical and rectangular shapes are shown in figure 3-8). The rectangular waveguide is used more frequently than the cylindrical waveguide.

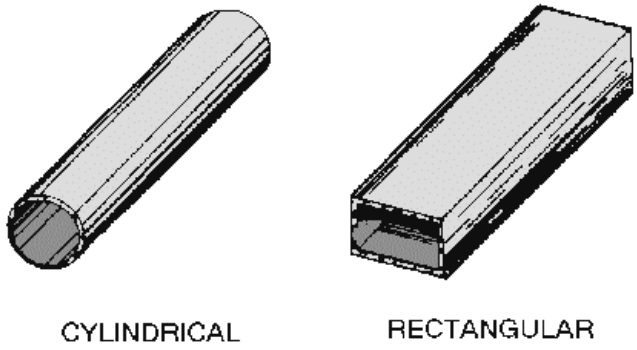


Figure 3-8.—Waveguides.

The term *waveguide* can be applied to all types of transmission lines in the sense that they are all used to guide energy from one point to another. However, usage has generally limited the term to mean a hollow metal tube or a dielectric transmission line. In this chapter, we use the term *waveguide* only to mean "hollow metal tube." It is interesting to note that the transmission of electromagnetic energy along a waveguide travels at a velocity somewhat slower than electromagnetic energy traveling through free space.

A waveguide may be classified according to its cross section (rectangular, elliptical, or circular), or according to the material used in its construction (metallic or dielectric). Dielectric waveguides are

seldom used because the dielectric losses for all known dielectric materials are too great to transfer the electric and magnetic fields efficiently.

The installation of a complete waveguide transmission system is somewhat more difficult than the installation of other types of transmission lines. The radius of bends in the waveguide must measure greater than two wavelengths at the operating frequency of the equipment to avoid excessive attenuation. The cross section must remain uniform around the bend. These requirements hamper installation in confined spaces. If the waveguide is dented, or if solder is permitted to run inside the joints, the attenuation of the line is greatly increased. Dents and obstructions in the waveguide also reduce its breakdown voltage, thus limiting the waveguide's power-handling capability because of possible arc over. Great care must be exercised during installation; one or two carelessly made joints can seriously inhibit the advantage of using the waveguide.

We will not consider the waveguide operation in this module, since waveguide theory is discussed in *NEETS*, Module 11, *Microwave Principles*.

Q4. List the five types of transmission lines in use today.

Q5. Name two of the three described uses of a two-wire open line.

Q6. What are the two primary disadvantages of a two-wire open line?

Q7. What type of transmission line is often used to connect a television set to its antenna?

Q8. What is the primary advantage of the shielded pair?

Q9. What are the two types of coaxial lines in use today?

Q10. What is the chief advantage of the air coaxial line?

Q11. List the three disadvantages of the air coaxial line.

Q12. List the two common types of waveguides in use today.

LOSSES IN TRANSMISSION LINES

The discussion of transmission lines so far has not directly addressed LINE LOSSES; actually some line losses occur in all lines. Line losses may be any of three types—COPPER, DIELECTRIC, and RADIATION or INDUCTION LOSSES.

NOTE: Transmission lines are sometimes referred to as rf lines. In this text the terms are used interchangeably.

Copper Losses

One type of copper loss is I^2R LOSS. In rf lines the resistance of the conductors is never equal to zero. Whenever current flows through one of these conductors, some energy is dissipated in the form of heat. This heat loss is a POWER LOSS. With copper braid, which has a resistance higher than solid tubing, this power loss is higher.

Another type of copper loss is due to SKIN EFFECT. When dc flows through a conductor, the movement of electrons through the conductor's cross section is uniform. The situation is somewhat different when ac is applied. The expanding and collapsing fields about each electron encircle other electrons. This phenomenon, called SELF INDUCTION, retards the movement of the encircled electrons.

The flux density at the center is so great that electron movement at this point is reduced. As frequency is increased, the opposition to the flow of current in the center of the wire increases. Current in the center of the wire becomes smaller and most of the electron flow is on the wire surface. When the frequency applied is 100 megahertz or higher, the electron movement in the center is so small that the center of the wire could be removed without any noticeable effect on current. You should be able to see that the effective cross-sectional area decreases as the frequency increases. Since resistance is inversely proportional to the cross-sectional area, the resistance will increase as the frequency is increased. Also, since power loss increases as resistance increases, power losses increase with an increase in frequency because of skin effect.

Copper losses can be minimized and conductivity increased in an rf line by plating the line with silver. Since silver is a better conductor than copper, most of the current will flow through the silver layer. The tubing then serves primarily as a mechanical support.

Dielectric Losses

DIELECTRIC LOSSES result from the heating effect on the dielectric material between the conductors. Power from the source is used in heating the dielectric. The heat produced is dissipated into the surrounding medium. When there is no potential difference between two conductors, the atoms in the dielectric material between them are normal and the orbits of the electrons are circular. When there is a potential difference between two conductors, the orbits of the electrons change. The excessive negative charge on one conductor repels electrons on the dielectric toward the positive conductor and thus distorts the orbits of the electrons. A change in the path of electrons requires more energy, introducing a power loss.

The atomic structure of rubber is more difficult to distort than the structure of some other dielectric materials. The atoms of materials, such as polyethylene, distort easily. Therefore, polyethylene is often used as a dielectric because less power is consumed when its electron orbits are distorted.

Radiation and Induction Losses

RADIATION and INDUCTION LOSSES are similar in that both are caused by the fields surrounding the conductors. Induction losses occur when the electromagnetic field about a conductor cuts through any nearby metallic object and a current is induced in that object. As a result, power is dissipated in the object and is lost.

Radiation losses occur because some magnetic lines of force about a conductor do not return to the conductor when the cycle alternates. These lines of force are projected into space as radiation and this results in power losses. That is, power is supplied by the source, but is not available to the load.

Q13. What are the three types of line losses associated with transmission lines?

Q14. Losses caused by skin effect and the I^2R (power) loss are classified as what type of loss?

Q15. What types of losses cause the dielectric material between the conductors to be heated?

LENGTH OF A TRANSMISSION LINE

A transmission line is considered to be electrically short when its physical length is short compared to a quarter-wavelength ($1/4\lambda$) of the energy it is to carry.

NOTE: In this module, for ease of reading, the value of the wavelength will be spelled out in some cases, and in other cases, the numerical value will be used.

A transmission line is electrically long when its physical length is long compared to a quarter-wavelength of the energy it is to carry. You must understand that the terms "short" and "long" are relative ones. For example, a line that has a physical length of 3 meters (approximately 10 feet) is considered quite short electrically if it transmits a radio frequency of 30 kilohertz. On the other hand, the same transmission line is considered electrically long if it transmits a frequency of 30,000 megahertz.

To show the difference in physical and electrical lengths of the lines mentioned above, compute the wavelength of the two frequencies, taking the 30-kilohertz example first:

Given:

$$\lambda = \frac{v}{f}$$

Where:

λ = Wavelength
v = Velocity of rf in free space
f = Frequency of transmission
Hz = Cycles per second

$$\lambda = \frac{300 \times 10^6 \text{ meters/second}}{30 \times 10^3 \text{ cycles/second (Hz)}}$$

$\lambda = 10 \times 10^3$ meters/cycle
λ = 10,000 meters, or approxiamtely 6 miles for complete wavelength

Now, computing the wavelength for the line carrying 30,000 megahertz:

$$\lambda = \frac{v}{f}$$

$$\lambda = \frac{300 \times 10^6 \text{ meters/second}}{30{,}000 \times 10^6 \text{ cycles/second (Hz)}}$$

$$\lambda = \frac{1}{100} \text{ meter/cycle}$$

λ = .01 meter, or approximately .03 foot for a complete wavelength

Thus, you can see that a 3-meter line is electrically very short for a frequency of 30 kilohertz. Also, the 3-meter line is electrically very long for a frequency of 30,000 megahertz.

When power is applied to a very short transmission line, practically all of it reaches the load at the output end of the line. This very short transmission line is usually considered to have practically no electrical properties of its own, except for a small amount of resistance.

However, the picture changes considerably when a long line is used. Since most transmission lines are electrically long (because of the distance from transmitter to antenna), the properties of such lines must be considered. Frequently, the voltage necessary to drive a current through a long line is considerably greater than the amount that can be accounted for by the impedance of the load in series with the resistance of the line.

TRANSMISSION LINE THEORY

The electrical characteristics of a two-wire transmission line depend primarily on the construction of the line. The two-wire line acts like a long capacitor. The change of its capacitive reactance is noticeable as the frequency applied to it is changed. Since the long conductors have a magnetic field about them when electrical energy is being passed through them, they also exhibit the properties of inductance. The values of inductance and capacitance presented depend on the various physical factors that we discussed earlier. For example, the type of line used, the dielectric in the line, and the length of the line must be considered. The effects of the inductive and capacitive reactances of the line depend on the frequency applied. Since no dielectric is perfect, electrons manage to move from one conductor to the other through the dielectric. Each type of two-wire transmission line also has a conductance value. This conductance value represents the value of the current flow that may be expected through the insulation. If the line is uniform (all values equal at each unit length), then one small section of the line may represent several feet. This illustration of a two-wire transmission line will be used throughout the discussion of transmission lines; but, keep in mind that the principles presented apply to all transmission lines. We will explain the theories using LUMPED CONSTANTS and DISTRIBUTED CONSTANTS to further simplify these principles.

LUMPED CONSTANTS

A transmission line has the properties of inductance, capacitance, and resistance just as the more conventional circuits have. Usually, however, the constants in conventional circuits are lumped into a single device or component. For example, a coil of wire has the property of inductance. When a certain amount of inductance is needed in a circuit, a coil of the proper dimensions is inserted. The inductance of the circuit is lumped into the one component. Two metal plates separated by a small space, can be used to supply the required capacitance for a circuit. In such a case, most of the capacitance of the circuit is lumped into this one component. Similarly, a fixed resistor can be used to supply a certain value of circuit resistance as a lumped sum. Ideally, a transmission line would also have its constants of inductance, capacitance, and resistance lumped together, as shown in figure 3-9. Unfortunately, this is not the case. Transmission line constants are *distributed*, as described below.

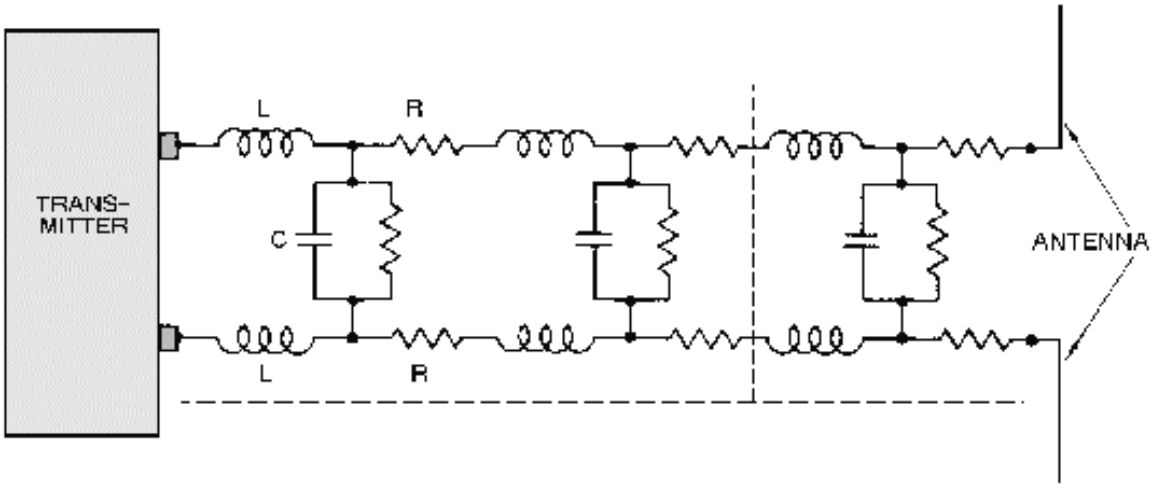


Figure 3-9.—Equivalent circuit of a two-wire transmission line.

DISTRIBUTED CONSTANTS

Transmission line constants, called *distributed constants*, are spread along the entire length of the transmission line and cannot be distinguished separately. The amount of inductance, capacitance, and resistance depends on the length of the line, the size of the conducting wires, the spacing between the wires, and the dielectric (air or insulating medium) between the wires. The following paragraphs will be useful to you as you study distributed constants on a transmission line.

Inductance of a Transmission Line

When current flows through a wire, magnetic lines of force are set up around the wire. As the current increases and decreases in amplitude, the field around the wire expands and collapses accordingly. The energy produced by the magnetic lines of force collapsing back into the wire tends to keep the current flowing in the same direction. This represents a certain amount of inductance, which is expressed in *microhenrys per unit length*. Figure 3-10 illustrates the inductance and magnetic fields of a transmission line.

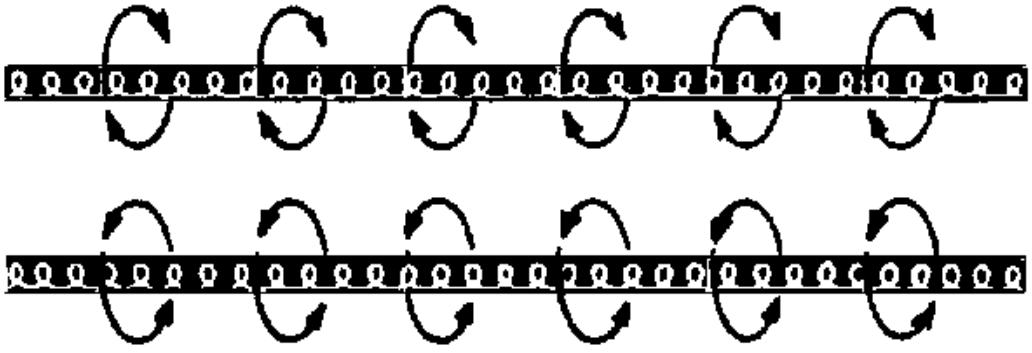


Figure 3-10.—Distributed inductance

Capacitance of a Transmission Line

Capacitance also exists between the transmission line wires, as illustrated in figure 3-11. Notice that the two parallel wires act as plates of a capacitor and that the air between them acts as a dielectric. The capacitance between the wires is usually expressed in *picofarads per unit length*. This electric field between the wires is similar to the field that exists between the two plates of a capacitor.

Figure 3-11.—Distributed capacitance.

Resistance of a Transmission Line

The transmission line shown in figure 3-12 has electrical resistance along its length. This resistance is usually expressed in *ohms per unit length* and is shown as existing continuously from one end of the line to the other.

Figure 3-12.—Distributed resistance.

Q16. What must the physical length of a transmission line be if it will be operated at 15,000,000 Hz? Use the formula:

$$\lambda = \frac{V}{f}$$

Q17. What are two of the three physical factors that determine the values of capacitance and inductance of a transmission line?

Q18. A transmission line is said to have distributed constants of inductance, capacitance, and resistance along the line. What units of measurement are used to express these constants?

Leakage Current

Since any dielectric, even air, is not a perfect insulator, a small current known as LEAKAGE CURRENT flows between the two wires. In effect, the insulator acts as a resistor, permitting current to pass between the two wires. Figure 3-13 shows this leakage path as resistors in parallel connected between the two lines. This property is called CONDUCTANCE (G) and is the opposite of resistance.

Conductance in transmission lines is expressed as the reciprocal of resistance and is usually given in *micromhos per unit length*.

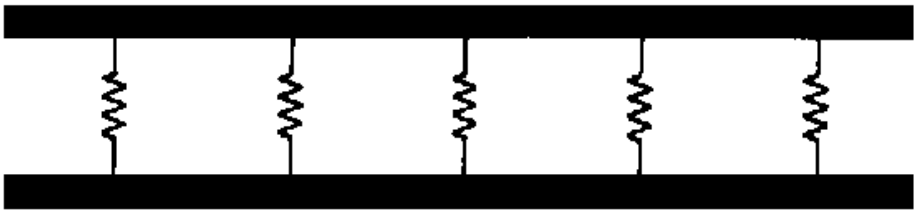


Figure 3-13.—Leakage in a transmission line.

ELECTROMAGNETIC FIELDS ABOUT A TRANSMISSION LINE

The distributed constants of resistance, inductance, and capacitance are basic properties common to all transmission lines and exist whether or not any current flow exists. As soon as current flow and voltage exist in a transmission line, another property becomes quite evident. This is the presence of an electromagnetic field, or lines of force, about the wires of the transmission line. The lines of force themselves are not visible; however, understanding the force that an electron experiences while in the field of these lines is very important to your understanding of energy transmission.

There are two kinds of fields; one is associated with voltage and the other with current. The field associated with voltage is called the ELECTRIC (E) FIELD. It exerts a force on any electric charge placed in it. The field associated with current is called a MAGNETIC (H) FIELD, because it tends to exert a force on any magnetic pole placed in it. Figure 3-14 illustrates the way in which the E fields and H fields tend to orient themselves between conductors of a typical two-wire transmission line. The illustration shows a cross section of the transmission lines. The E field is represented by solid lines and the H field by dotted lines. The arrows indicate the direction of the lines of force. Both fields normally exist together and are spoken of collectively as the electromagnetic field.

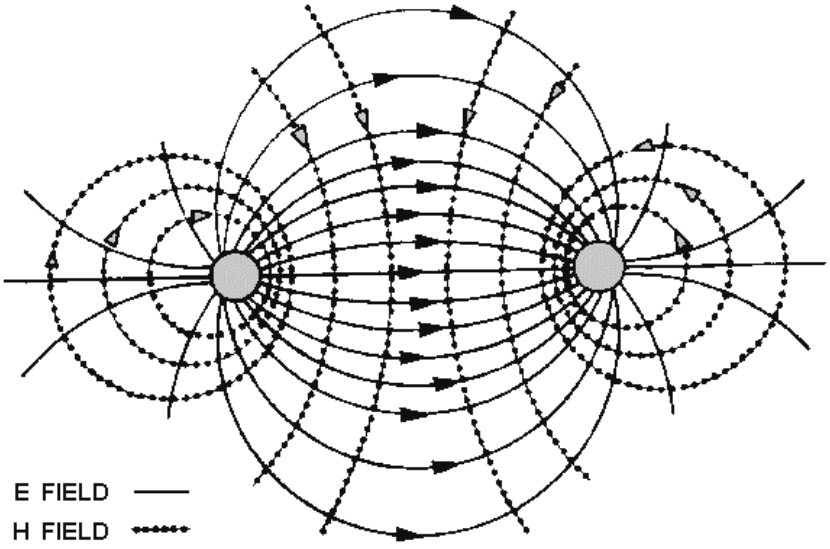


Figure 3-14.—Fields between conductors.

CHARACTERISTIC IMPEDANCE OF A TRANSMISSION LINE

You learned earlier that the maximum (and most efficient) transfer of electrical energy takes place when the source impedance is matched to the load impedance. This fact is very important in the study of transmission lines and antennas. If the characteristic impedance of the transmission line and the load impedance are equal, energy from the transmitter will travel down the transmission line to the antenna with no power loss caused by reflection.

Definition and Symbols

Every transmission line possesses a certain CHARACTERISTIC IMPEDANCE, usually designated as Z_0. Z_0 is the ratio of E to I at every point along the line. If a load equal to the characteristic impedance is placed at the output end of any length of line, the same impedance will appear at the input terminals of the line. The characteristic impedance is the only value of impedance for any given type and size of line that acts in this way. The characteristic impedance determines the amount of current that can flow when a given voltage is applied to an infinitely long line. Characteristic impedance is comparable to the resistance that determines the amount of current that flows in a dc circuit.

In a previous discussion, lumped and distributed constants were explained. Figure 3-15, view A, shows the properties of resistance, inductance, capacitance, and conductance combined in a short section of two-wire transmission line. The illustration shows the evenly distributed capacitance as a single lumped capacitor and the distributed conductance as a lumped leakage path. Lumped values may be used for transmission line calculations if the physical length of the line is very short compared to the wavelength of energy being transmitted. Figure 3-15, view B, shows all four properties lumped together and represented by their conventional symbols.

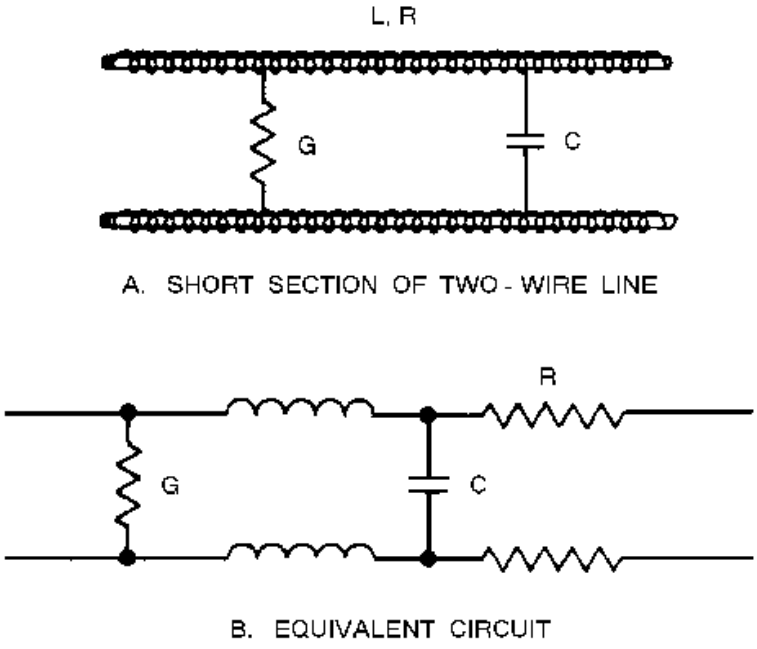


Figure 3-15.—Short section of two-wire transmission line and equivalent circuit.

Q19. Describe the leakage current in a transmission line and in what unit it is expressed.

Q20. All the power sent down a transmission line from a transmitter can be transferred to an antenna under what optimum conditions?

Q21. What symbol is used to designate the characteristic impedance of a line, and what two variables does it compare?

Characteristic Impedance and the Infinite Line

Several short sections, as shown in figure 3-15, can be combined to form a large transmission line, as shown in figure 3-16. Current will flow if voltage is applied across points K and L. In fact, any circuit, such as that represented in figure 3-16, view A, has a certain current flow for each value of applied voltage. The ratio of the voltage to the current is the impedance (Z).

Recall that:

$$Z = \frac{E}{I}$$

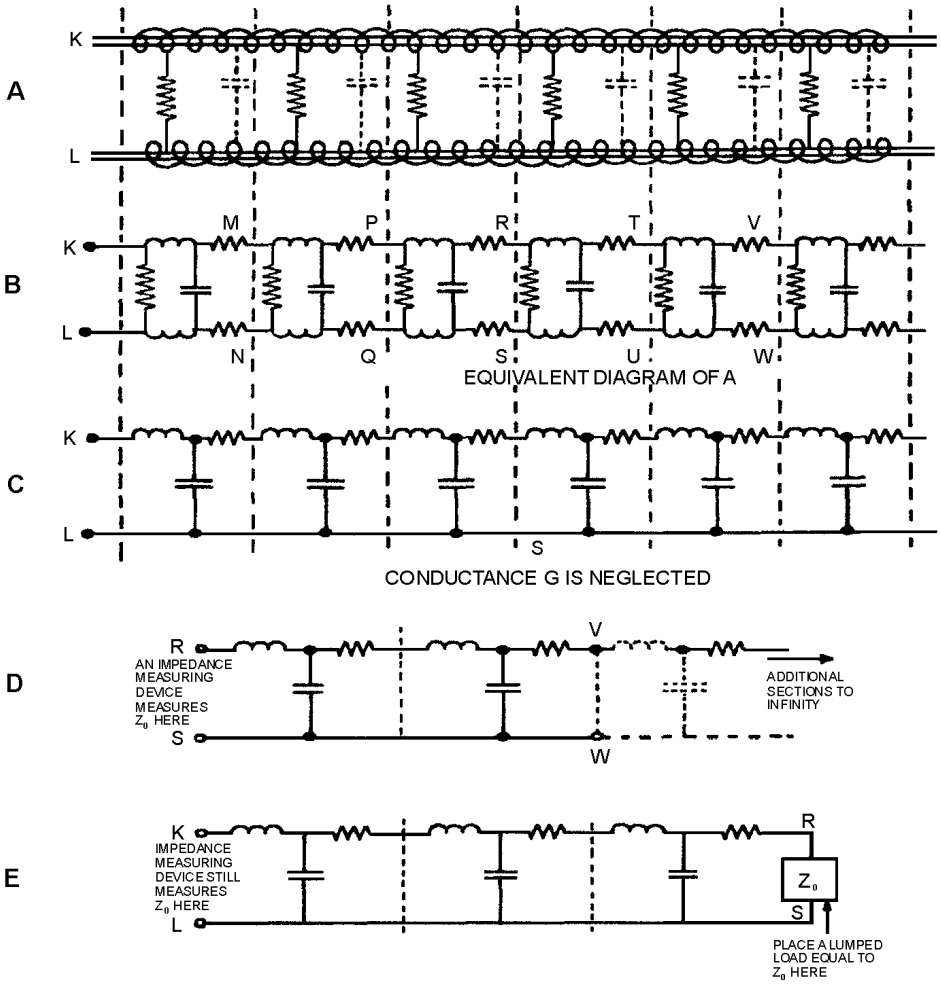


Figure 3-16.—Characteristic impedance.

The impedance presented to the input terminals of the transmission line is not merely the resistance of the wire in series with the impedance of the load. The effects of series inductance and shunt capacitance of the line itself may overshadow the resistance, and even the load, as far as the input terminals are concerned.

To find the input impedance of a transmission line, determine the impedance of a single section of line. The impedance between points K and L, in view B of figure 3-16, can be calculated by the use of series-parallel impedance formulas, provided the impedance across points M and N is known. But since this section is merely one small part of a longer line, another similar section is connected to points M and N. Again, the impedance across points K and L of the two sections can be calculated, provided the impedance of the third section is known. This process of adding one section to another can be repeated endlessly. The addition of each section produces an impedance across points K and L of a new and lower value. However, after many sections have been added, each successive added section has less and less effect on the impedance across points K and L. If sections are added to the line endlessly, the line is infinitely long, and a certain finite value of impedance across points K and L is finally reached.

In this discussion of transmission lines, the effect of conductance (G) is minor compared to that of inductance (L) and capacitance (C), and is frequently neglected. In figure 3-16, view C, G is omitted and the inductance and resistance of each line can be considered as one line.

Let us assume that the sections of view C continue to the right with an infinite number of sections. When an infinite number of sections extends to the right, the impedance appearing across K and L is Z_0. If the line is cut at R and S, an infinite number of sections still extends to the right since the line is endless in that direction. Therefore, the impedance now appearing across points R and S is also Z_0, as illustrated in view D. You can see that if only the first three sections are taken and a load impedance of Z_0 is connected across points R and S, the impedance across the input terminals K and L is still Z_0. The line continues to act as an infinite line. This is illustrated in view E.

Figure 3-17, view A, illustrates how the characteristic impedance of an infinite line can be calculated. Resistors are added in series parallel across terminals K and L in eight steps, and the resultant impedances are noted. In step 1 the impedance is infinite; in step 2 the impedance is 110 ohms. In step 3 the impedance becomes 62.1 ohms, a change of 47.9 ohms. In step 4 the impedance is 48.5 ohms, a change of only 13.6 ohms. The resultant changes in impedance from each additional increment become progressively smaller. Eventually, practically no change in impedance results from further additions to the line. The total impedance of the line at this point is said to be at its characteristic impedance; which, in this case, is 37 ohms. This means that an infinite line constructed as indicated in step 8 could be effectively replaced by a 37-ohm resistor. View B shows a 37-ohm resistor placed in the line at various points to replace the infinite line of step 8 in view A. There is no change in total impedance.

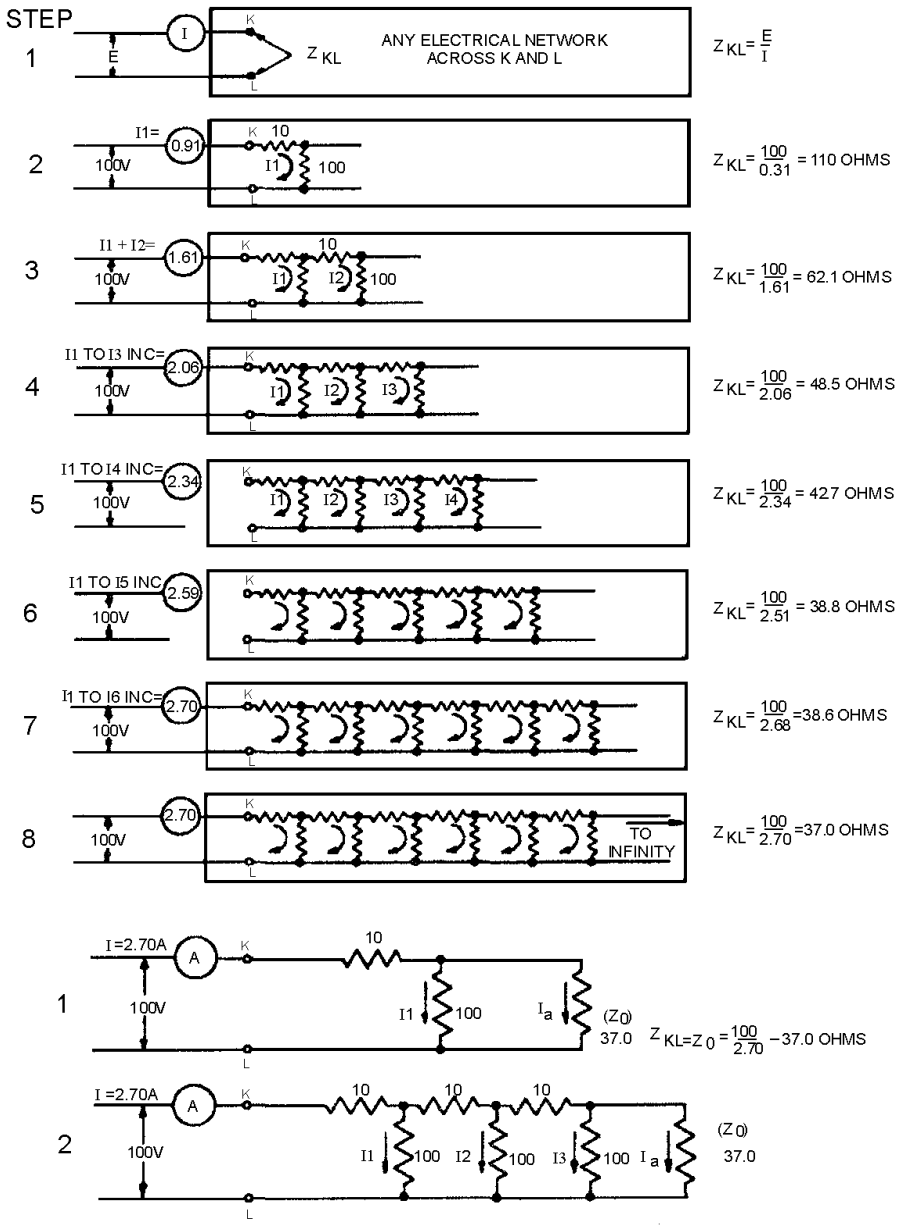


Figure 3-17.—Termination of a line.

In figure 3-17, resistors were used to show impedance characteristics for the sake of simplicity. Figuring the actual impedance of a line having reactance is very similar, with inductance taking the place of the series resistors and capacitance taking the place of the shunt resistors. The characteristic impedance of lines in actual use normally lies between 50 and 600 ohms.

When a transmission line is "short" compared to the length of the radio-frequency waves it carries, the opposition presented to the input terminals is determined primarily by the load impedance. A small amount of power is dissipated in overcoming the resistance of the line. However, when the line is "long" and the load is an incorrect impedance, the voltages necessary to drive a given amount of current through the line cannot be accounted for by considering just the impedance of the load in series with the

impedance of the line. The line has properties other than resistance that affect input impedance. These properties are inductance in series with the line, capacitance across the line, resistance leakage paths across the line, and certain radiation losses.

Q22. What is the range of the characteristic impedance of lines used in actual practice?

VOLTAGE CHANGE ALONG A TRANSMISSION LINE

Let us summarize what we have just discussed. In an electric circuit, energy is stored in electric and magnetic fields. These fields must be brought to the load to transmit that energy. At the load, energy contained in the fields is converted to the desired form of energy.

Transmission of Energy

When the load is connected directly to the source of energy, or when the transmission line is short, problems concerning current and voltage can be solved by applying Ohm's law. When the transmission line becomes long enough so the time difference between a change occurring at the generator and the change appearing at the load becomes appreciable, analysis of the transmission line becomes important.

Dc Applied to a Transmission Line

In figure 3-18, a battery is connected through a relatively long two-wire transmission line to a load at the far end of the line. At the instant the switch is closed, neither current nor voltage exists on the line. When the switch is closed, point A becomes a positive potential, and point B becomes negative. These points of difference in potential move down the line. However, as the initial points of potential leave points A and B, they are followed by new points of difference in potential which the battery adds at A and B. This is merely saying that the battery maintains a constant potential difference between points A and B. A short time after the switch is closed, the initial points of difference in potential have reached points A' and B'; the wire sections from points A to A' and points B to B' are at the same potential as A and B, respectively. The points of charge are represented by plus (+) and minus (-) signs along the wires. The directions of the currents in the wires are represented by the arrowheads on the line, and the direction of travel is indicated by an arrow below the line. Conventional lines of force represent the electric field that exists between the opposite kinds of charge on the wire sections from A to A' and B to B'. Crosses (tails of arrows) indicate the magnetic field created by the electric field moving down the line. The moving electric field and the accompanying magnetic field constitute an electromagnetic wave that is moving from the generator (battery) toward the load. This wave travels at approximately the speed of light in free space. The energy reaching the load is equal to that developed at the battery (assuming there are no losses in the transmission line). If the load absorbs all of the energy, the current and voltage will be evenly distributed along the line.

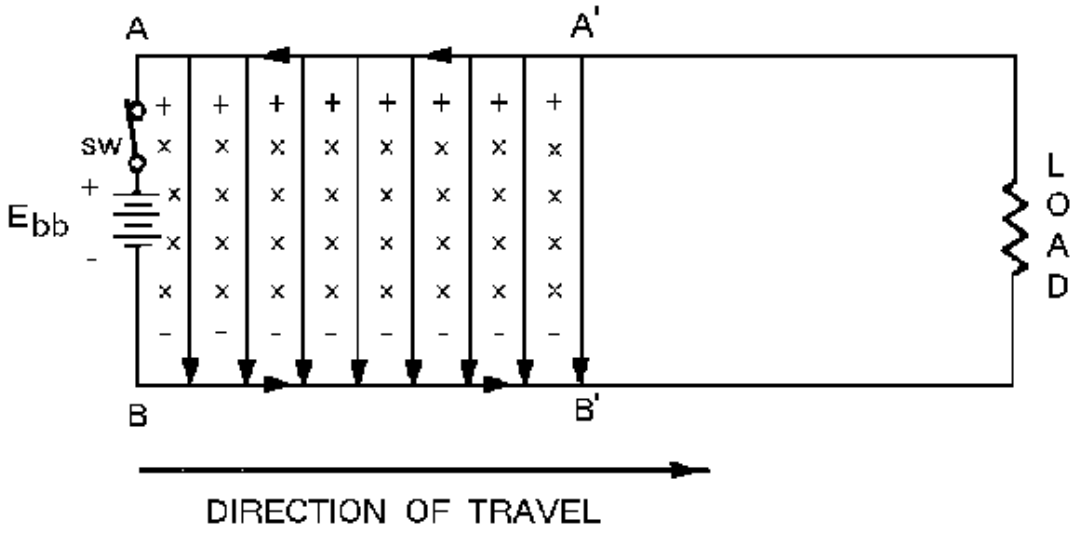


Figure 3-18.—Dc voltage applied to a line.

Ac Applied to a Transmission Line

When the battery of figure 3-18 is replaced by an ac generator (fig. 3-19), each successive instantaneous value of the generator voltage is propagated down the line at the speed of light. The action is similar to the wave created by the battery except that the applied voltage is sinusoidal instead of constant. Assume that the switch is closed at the moment the generator voltage is passing through zero and that the next half cycle makes point A positive. At the end of one cycle of generator voltage, the current and voltage distribution will be as shown in figure 3-19.

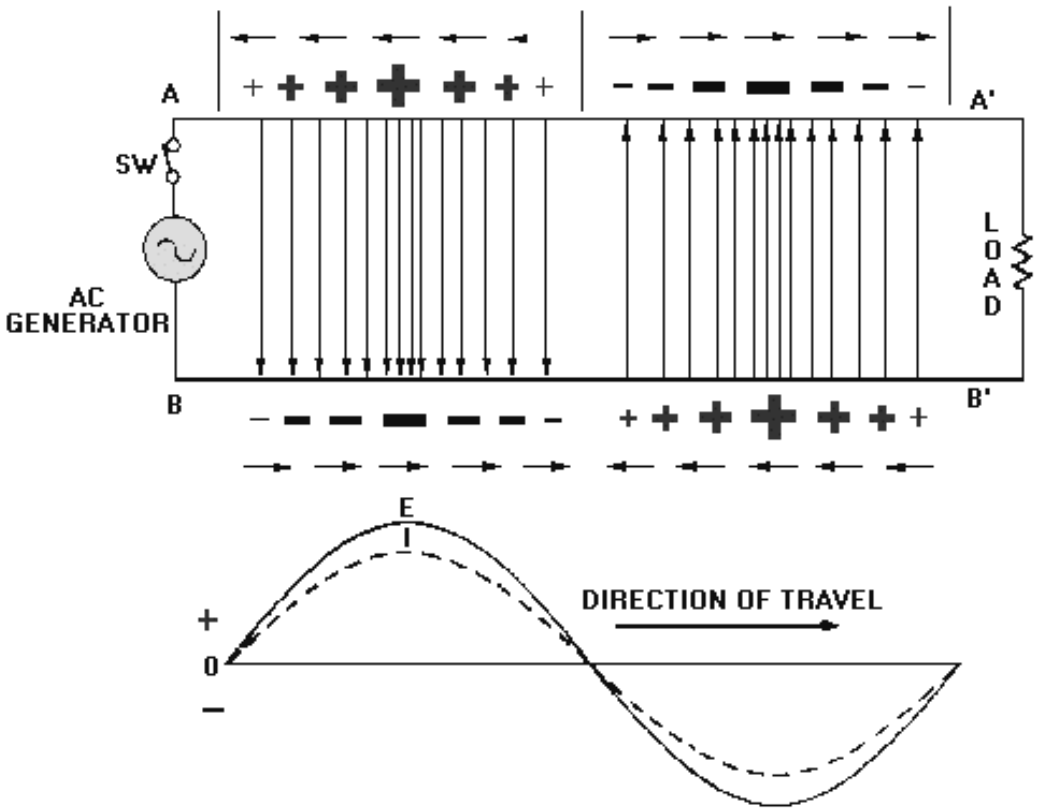


Figure 3-19.—Ac voltage applied to a line.

In this illustration the conventional lines of force represent the electric fields. For simplicity, the magnetic fields are not shown. Points of charge are indicated by plus (+) and minus (–) signs, the larger signs indicating points of higher amplitude of both voltage and current. Short arrows indicate direction of current (electron flow). The waveform drawn below the transmission line represents the voltage (E) and current (I) waves. The line is assumed to be infinite in length so there is no reflection. Thus, traveling sinusoidal voltage and current waves continually travel in phase from the generator toward the load, or far end of the line. Waves traveling from the generator to the load are called INCIDENT WAVES. Waves traveling from the load back to the generator are called REFLECTED WAVES and will be explained in later paragraphs.

Dc Applied to an Infinite Line

Figure 3-20 shows a battery connected to a circuit that is the equivalent of a transmission line. In this line the series resistance and shunt conductance are not shown. In the following discussion the line will be considered to have no losses.

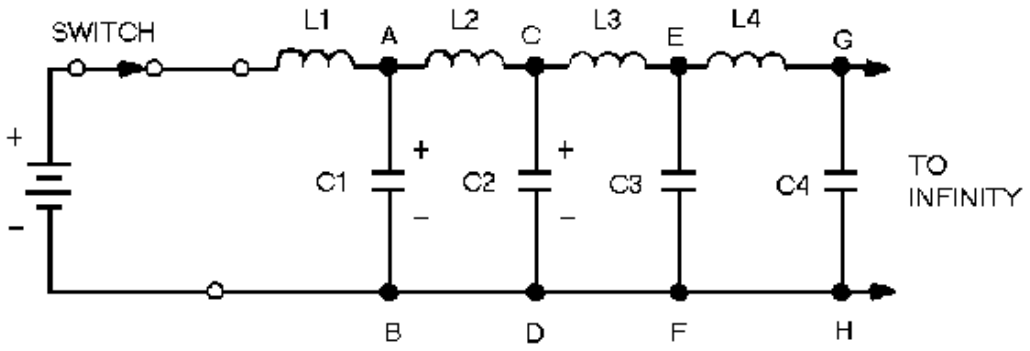


Figure 3-20.—Dc applied to an equivalent transmission line.

As the switch is closed, the battery voltage is applied to the input terminals of the line. Now, C1 has no charge and appears, effectively, as a short circuit across points A and B. The full battery voltage appears across inductor L1. Inductor L1 opposes the change of current (0 now) and limits the rate of charge of C1.

Capacitor C2 cannot begin to charge until after C1 has charged. No current can flow beyond points A and B until C1 has acquired some charge. As the voltage across C1 increases, current through L2 and C2 charges C2. This action continues down the line and charges each capacitor, in turn, to the battery voltage. Thus a voltage wave is traveling along the line. Beyond the wavefront, the line is uncharged. Since the line is infinitely long, there will always be more capacitors to be charged, and current will not stop flowing. Thus current will flow indefinitely in the line.

Notice that current flows to charge the capacitors along the line. The flow of current is not advanced along the line until a voltage is developed across each preceding capacitor. In this manner voltage and current move down the line together in phase.

Ac Applied to an Infinite Line

An rf line displays similar characteristics when an ac voltage is applied to its sending end or input terminals. In figure 3-21, view A, an ac voltage is applied to the line represented by the circuit shown.

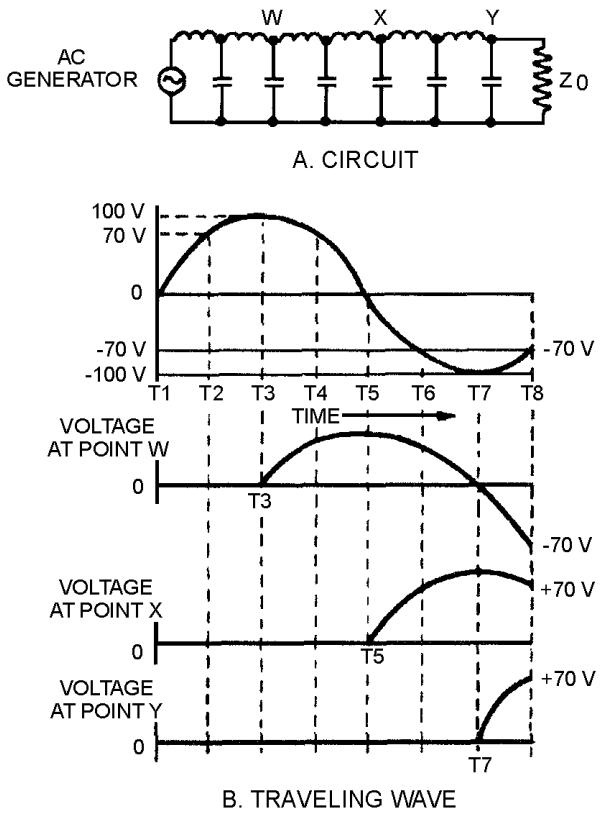


Figure 3-21.—Ac applied to an equivalent transmission line.

In view B the generator voltage starts from zero (T1) and produces the voltage shown. As soon as a small voltage change is produced, it starts its journey down the line while the generator continues to produce new voltages along a sine curve. At T2 the generator voltage is 70 volts. The voltages still move along the line until, at T3, the first small change arrives at point W, and the voltage at that point starts increasing. At T5, the same voltage arrives at point X on the line. Finally, at T7, the first small change arrives at the receiving end of the line. Meanwhile, all the changes in the sine wave produced by the generator pass each point in turn. The amount of time required for the changes to travel the length of the line is the same as that required for a dc voltage to travel the same distance.

At T7, the voltage at the various points on the line is as follows:

At the generator:	-100 V
At point W:	0 V
At point X:	+100 V
At point Y:	0 V

If these voltages are plotted along the length of the line, the resulting curve is like the one shown in figure 3-22, view A. Note that such a curve of instantaneous voltages resembles a sine wave. The changes in voltage that occur between T7 and T8 are as follows:

At the generator:	Rise from	-100 V to -70 V
At point W:	Drop from	0 V to -70 V
At point X:	Drop from	+100 V to +70 V
At point Y:	Rise from	0 V to + 70 V

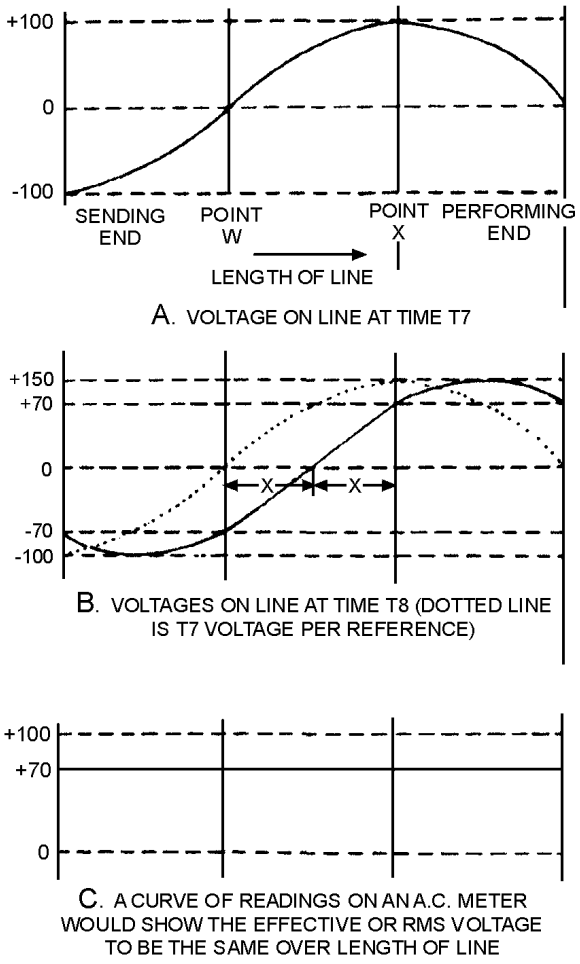


Figure 3-22.—Instantaneous voltages along a transmission line.

A plot of these new voltages produces the solid curve shown in figure 3-22, view B. For reference, the curve from T7 is drawn as a dotted line. The solid curve has exactly the same shape as the dotted curve, but has moved to the right by the distance X. Another plot at T9 would show a new curve similar to the one at T8, but moved to the right by the distance Y.

By analyzing the points along the graph just discussed, you should be able to see that the actions associated with voltage changes along an rf line are as follows:

1. All instantaneous voltages of the sine wave produced by the generator travel down the line in the order they are produced.

2. At any point, a sine wave can be obtained if all the instantaneous voltages passing the point are plotted. An oscilloscope can be used to plot these values of instantaneous voltages against time.

3. The instantaneous voltages (oscilloscope displays) are the same in all cases except that a phase difference exists in the displays seen at different points along the line. The phase changes continually with respect to the generator until the change is 360 degrees over a certain length of line.

4. All parts of a sine wave pass every point along the line. A plot of the readings of an ac meter (which reads the effective value of the voltage over a given time) taken at different points along the line shows that the voltage is constant at all points. This is shown in view C of figure 3-22.

5. Since the line is terminated with a resistance equal to Z_0, the energy arriving at the end of the line is absorbed by the resistance.

VELOCITY OF WAVE PROPAGATION

If a voltage is initially applied to the sending end of a line, that same voltage will appear later some distance from the sending end. This is true regardless of any change in voltage, whether the change is a jump from zero to some value or a drop from some value to zero. The voltage change will be conducted down the line at a constant rate.

Recall that the inductance of a line delays the charging of the line capacitance. The velocity of propagation is therefore related to the values of L and C. If the inductance and capacitance of the rf line are known, the time required for any waveform to travel the length of the line can be determined. To see how this works, observe the following relationship:

$$Q = IT$$

This formula shows that the total charge or quantity is equal to the current multiplied by the time the current flows. Also:

$$Q = CE$$

This formula shows that the total charge on a capacitor is equal to the capacitance multiplied by the voltage across the capacitor.

If the switch in figure 3-23 is closed for a given time, the quantity (Q) of electricity leaving the battery can be computed by using the equation Q = IT. The electricity leaves the battery and goes into the line, where a charge is built up on the capacitors. The amount of this charge is computed by using the equation Q = CE.

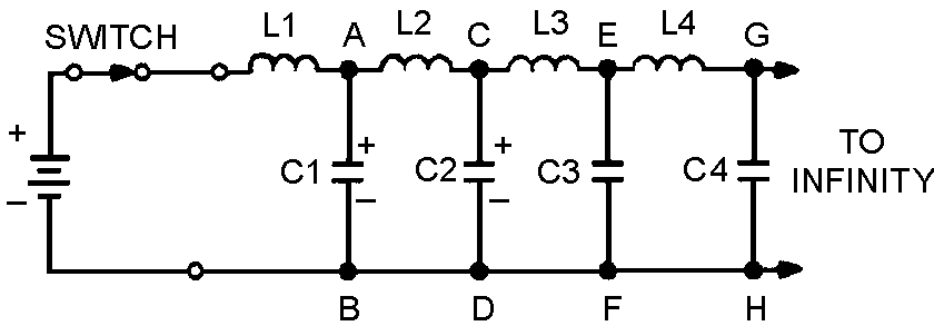


Figure 3-23.—Dc applied to an equivalent transmission line.

Since none of the charge is lost, the total charge leaving the battery during T is equal to the total charge on the line. Therefore:

$$Q = IT = CE$$

As each capacitor accumulates a charge equal to CE, the voltage across each inductor must change. As C1 in figure 3-23 charges to a voltage of E, point A rises to a potential of E volts while point B is still at zero volts. This makes E appear across L2. As C2 charges, point B rises to a potential of E volts as did point A. At this time, point B is at E volts and point C rises. Thus, we have a continuing action of voltage moving down the infinite line.

In an inductor, these circuit components are related, as shown in the formula

$$E = L\left(\frac{\Delta I}{\Delta T}\right).$$

This shows that the voltage across the inductor is directly proportional to inductance and the change in current, but inversely proportional to a change in time. Since current and time start from zero, the change in time (ΔT) and the change in current (ΔI) are equal to the final time (T) and final current (I). For this case the equation becomes:

$$ET = LI$$

If voltage E is applied for time (T) across the inductor (L), the final current (I) will flow. The following equations show how the three terms (T, L, and C) are related:

$$IT = CE$$
$$ET = LI$$

For convenience, you can find T in terms of L and C in the following manner. Multiply the left and right member of each equation as follows:

$$(IT)(ET) = (CE)(LI)$$

Then: $EIT^2 = LCEI$

Dividing by (EI): $T^2 = LC$

and $T = \sqrt{LC}$

This final equation is used for finding the time required for a voltage change to travel a unit length, since L and C are given in terms of unit length. The velocity of the waves may be found by:

$$V = \frac{D}{T} \text{ or } V = \frac{D}{\sqrt{LC}}$$

Where: D is the physical length of a unit

This is the rate at which the wave travels over a unit length. The units of L and C are henrys and farads, respectively. T is in seconds per unit length and V is in unit lengths per second.

DETERMINING CHARACTERISTIC IMPEDANCE

As previously discussed, an infinite transmission line exhibits a definite input impedance. This impedance is the CHARACTERISTIC IMPEDANCE and is independent of line length. The exact value of this impedance is the ratio of the input voltage to the input current. If the line is infinite or is terminated in a resistance equal to the characteristic impedance, voltage and current waves traveling the line are in phase. To determine the characteristic impedance or voltage-to-current ratio, use the following procedure:

Divide the equation:

$ET = LI$ by $IT = CE$

$$\frac{ET}{IT} = \frac{LI}{CE}$$

Multiply by $\frac{E}{I}$:

$$\frac{E^2T}{I^2T} = \frac{LIE}{CEI}$$

Simplify:

$$\frac{E^2}{I^2} = \frac{L}{C}$$

Take the square root:

$$\frac{E}{I} = \sqrt{\frac{L}{C}} = Z_0 \text{(characteristic impedance)}$$

Example:

A problem using this equation will illustrate how to determine the characteristics of a transmission line. Assume that the line shown in figure 3-23 is 1000 feet long. A 100-foot (approximately 30.5 meter) section is measured to determine L and C. The section is found to have an inductance of 0.25 millihenries and a capacitance of 1000 picofarads. Find the characteristic impedance of the line and the velocity of the wave on the line.

The characteristic impedance is:

$$Z_0 = \sqrt{LC}$$

$$Z_0 = \sqrt{\frac{0.25 \times 10^{-3}}{1000 \times 10^{-12}}}$$

$$Z_0 = \sqrt{0.25 \times 10^{6}}$$

$$Z_0 = 0.5 \times 10^{3}$$

$$Z_0 = 500\Omega$$

If any other unit length had been considered, the values of L and C would be different, but their ratio would remain the same as would the characteristic impedance.

The formula for T is:

$$T = \sqrt{LC}$$

$$T = \sqrt{0.25 \times 10^{-3} \times 1000 \times 10^{-12}}$$

$$T = \sqrt{.025 \times 10^{-12}}$$

$$T = 0.5 \times 10^{-6} \text{ second}$$

$$T = 0.5 \text{ microsecond}$$

The formula for the velocity of a wave is:

$$V = \frac{D}{T}$$

$$V = \frac{100 \text{ feet}}{0.5 \times 10^{-6} \text{ second}}$$

$$V = 200 \times 10^{6} \text{ feet/second}$$

$$V = 200{,}000{,}000 \text{ feet/second}$$

REFLECTIONS ON A TRANSMISSION LINE

Transmission line characteristics are based on an infinite line. A line cannot always be terminated in its characteristic impedance since it is sometimes operated as an OPEN-ENDED line and other times as a SHORT-CIRCUIT at the receiving end. If the line is open-ended, it has a terminating impedance that is infinitely large. If a line is not terminated in characteristic impedance, it is said to be finite.

When a line is not terminated in Z_0, the incident energy is not absorbed but is returned along the only path available—the transmission line. Thus, the behavior of a finite line may be quite different from that of the infinite line.

REFLECTION OF DC VOLTAGE FROM AN OPEN CIRCUIT

The equivalent circuit of an open-ended transmission line is shown in figure 3-24, view A. Again, losses are to be considered as negligible, and L is lumped in one branch. Assume that (1) the battery in this circuit has an internal impedance equal to the characteristic impedance of the transmission line ($Z_i = Z_0$); (2) the capacitors in the line are not charged before the battery is connected; and (3) since the line is open-ended, the terminating impedance is infinitely large.

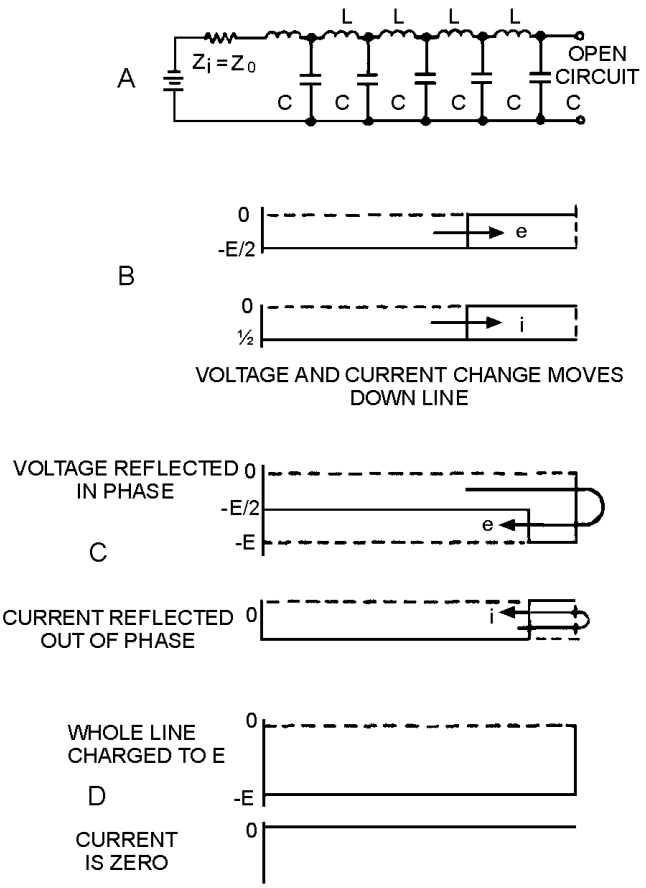


Figure 3-24.—Reflection from an open-ended line.

When the battery is connected to the sending end as shown, a negative voltage moves down the line. This voltage charges each capacitor, in turn, through the preceding inductor. Since Z_i equals Z_0, one-half the applied voltage will appear across the internal battery impedance, Z_i, and one-half across the impedance of the line, Z_0. Each capacitor is then charged to E/2 (view B). When the last capacitor in the line is charged, there is no voltage across the last inductor and current flow through the last inductor stops. With no current flow to maintain it, the magnetic field in the last inductor collapses and forces current to continue to flow in the same direction into the last capacitor. Because the direction of current has not changed, the capacitor charges in the same direction, thereby increasing the charge in the capacitor. Since the energy in the magnetic field equals the energy in the capacitor, the energy transfer to the capacitor doubles the voltage across the capacitor. The last capacitor is now charged to E volts and the current in the last inductor drops to zero.

At this point, the same process takes place with the next to the last inductor and capacitor. When the magnetic field about the inductor collapses, current continues to flow into the next to the last capacitor, charging it to E volts. This action continues backward down the line until the first capacitor has been fully charged to the applied voltage. This change of voltage, moving backward down the line, can be thought of in the following manner. The voltage, arriving at the end of the line, finds no place to go and returns to the sending end with the same polarity (view C). Such action is called REFLECTION.

When a reflection of voltage occurs on an open-ended line, the polarity is unchanged. The voltage change moves back to the source, charging each capacitor in turn until the first capacitor is charged to the

source voltage and the action stops (view D). As each capacitor is charged, current in each inductor drops to zero, effectively reflecting the current with the opposite polarity (view C). Reflected current of opposite polarity cancels the original current at each point, and the current drops to zero at that point. When the last capacitor is charged, the current from the source stops flowing (view D).

Important facts to remember in the reflection of dc voltages in open-ended lines are:

- Voltage is reflected from an open end without change in polarity, amplitude, or shape.
- Current is reflected from an open end with opposite polarity and without change in amplitude or shape.

REFLECTION OF DC VOLTAGE FROM A SHORT CIRCUIT

A SHORT-CIRCUITED line affects voltage change differently from the way an open-circuited line affects it. The voltage across a perfect short circuit must be zero; therefore, no power can be absorbed in the short, and the energy is reflected toward the generator.

The initial circuit is shown in figure 3-25, view A. The initial voltage and current waves (view B) are the same as those given for an infinite line. In a short-circuited line the voltage change arrives at the last inductor in the same manner as the waves on an open-ended line. In this case, however, there is no capacitor to charge. The current through the final inductor produces a voltage with the polarity shown in view C. When the field collapses, the inductor acts as a battery and forces current through the capacitor in the opposite direction, causing it to discharge (view D). Since the amount of energy stored in the magnetic field is the same as that in the capacitor, the capacitor discharges to zero.

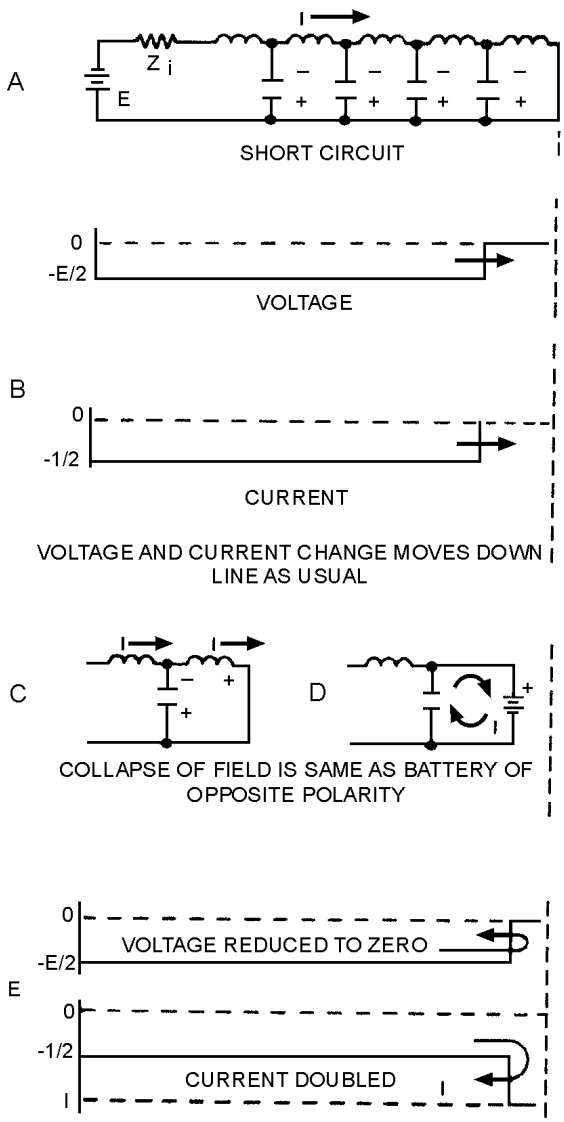


Figure 3-25.—Reflection from a short-circuited line.

Now there is no voltage to maintain the current through the next to the last inductor. Therefore, this inductor discharges the next to the last capacitor.

As each capacitor is discharged to zero, the next inductor effectively becomes a new source of voltage. The amplitude of each of these voltages is equal to E/2, but the polarity is the opposite of the battery at the input end of the line. The collapsing field around each inductor, in turn, produces a voltage that forces the current to continue flowing in the same direction, adding to the current from the source to make it 2I. This action continues until all the capacitors are discharged (view E).

Reflected waves from a short-circuited transmission line are characterized as follows:

- The reflected voltage has the opposite polarity but the same amplitude as the incident wave.
- The reflected current has the same polarity and the same amplitude as the incident current.

REFLECTION OF AC VOLTAGE FROM AN OPEN CIRCUIT

In most cases where rf lines are used, the voltages applied to the sending end are ac voltages. The action at the receiving end of the line is exactly the same for ac as for dc. In the open-ended line, shown in figure 3-26, view A, the generated ac voltage is distributed along the line, shown in view B. This voltage is distributed in such a way that as each instantaneous voltage arrives at the end, it is reflected with the same polarity and amplitude. When ac is used, this reflection is in phase. Each of the reflected voltages travels back along the line until it reaches the generator. If the generator impedance is the same as the line impedance, energy arriving at the generator is absorbed and not reflected again. Now two voltages are on the line.

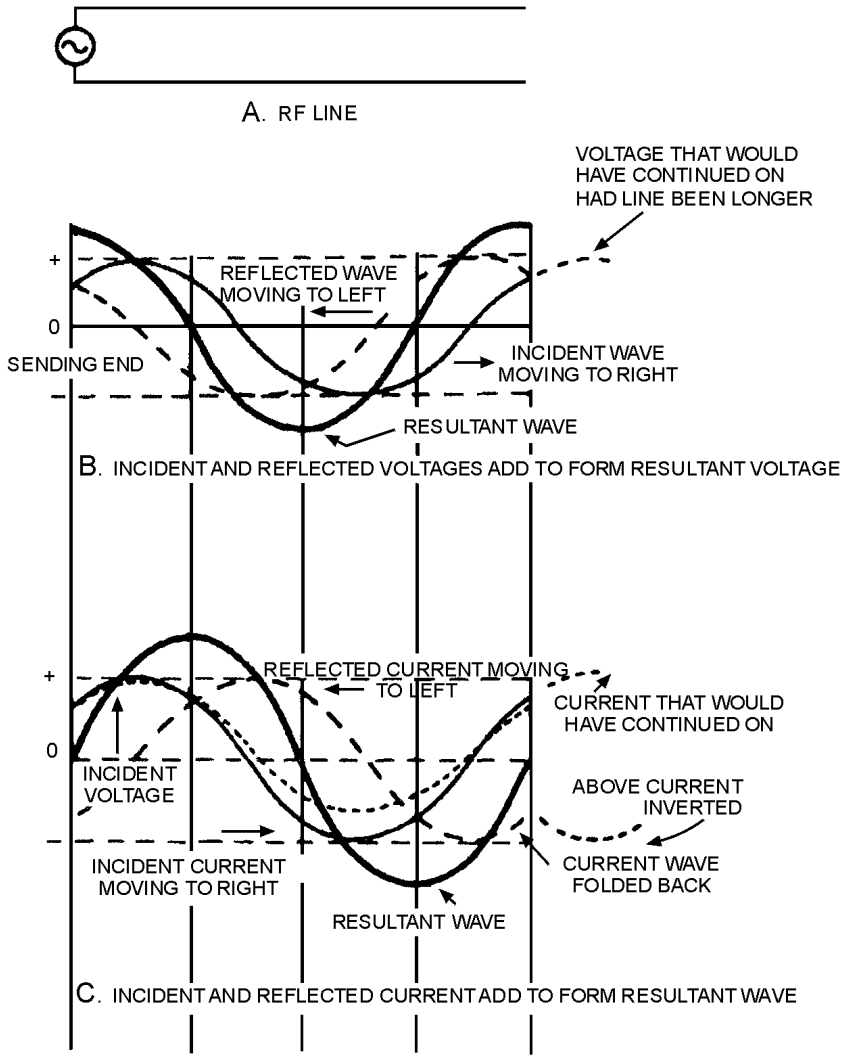


Figure 3-26.—Formation of standing waves.

View B shows how two waves of the same frequency and amplitude moving in opposite directions on the same conductor will combine to form a resultant wave. The small solid line is moving steadily from left to right and is the INCIDENT WAVE (from the source). The broken-line waveform is moving from right to left and is the REFLECTED WAVE. The resultant waveform, the heavy line, is found by algebraically adding instantaneous values of the two waveforms. The resultant waveform has an

instantaneous peak amplitude that is equal to the sum of the peak amplitudes of the incident and reflected waves. Since most indicating instruments are unable to separate these voltages, they show the vector sum. An oscilloscope is usually used to study the instantaneous voltages on rf lines.

Since two waves of voltage are moving on the line, you need to know how to distinguish between the two. The voltages moving toward the receiving end are called INCIDENT VOLTAGES, and the whole waveshape is called the INCIDENT WAVE. The wave moving back to the sending end after reflection is called the REFLECTED WAVE. The resultant voltage curve (view B of figure 3-26) shows that the voltage is maximum at the end of the line, a condition that occurs across an open circuit.

Another step in investigating the open-circuited rf line is to see how the current waves act. The incident current wave is the solid line in figure 3-26, view C. The voltage is represented by the dotted line. The current is in phase with the voltage while traveling toward the receiving end. At the end of the line, the current is reflected in the opposite polarity; that is, it is shifted 180 degrees in phase, but its amplitude remains the same. The reflected wave of current is shown by dashed lines in view C. The heavy-line curve represents the sum of the two instantaneous currents and is the resultant wave. Notice that current is zero at the end of the line. This is reasonable, since there can be no current flow through an open circuit.

Views B and C of figure 3-26 show the voltage and current distribution along a transmission line at a point about 1/8 after a maximum voltage or current reaches the end of the line. Since the instantaneous values are continuously changing during the generation of a complete cycle, a large number of these pictures are required to show the many different relationships.

Figure 3-27 shows the incident and reflected waveshapes at several different times. The diagrams in the left column of figure 3-27 (representing *voltage*) show the incident wave and its reflection without change in polarity. In figure 3-27, waveform (1), the incident wave and the reflected wave are added algebraically to produce the resultant wave indicated by the heavy line. In waveform (2), a zero point preceding the negative-going cycle of the incident wave is at the end of the line. The reflected wave and incident wave are 180 degrees out of phase at all points. (The reflected wave is the positive cycle that just preceded the negative cycle now approaching the end of the line.) The resultant of the incident and reflected waves is zero at all points along the line. In waveform (3), the waves have moved $1/8\lambda$ along the line; the incident wave has moved 45 degrees to the right, and the reflected wave has moved 45 degrees to the left. The resultant voltage, shown by the heavy line, has a maximum negative at the end of the line and a maximum positive $1/2\lambda$ from the end of the line.

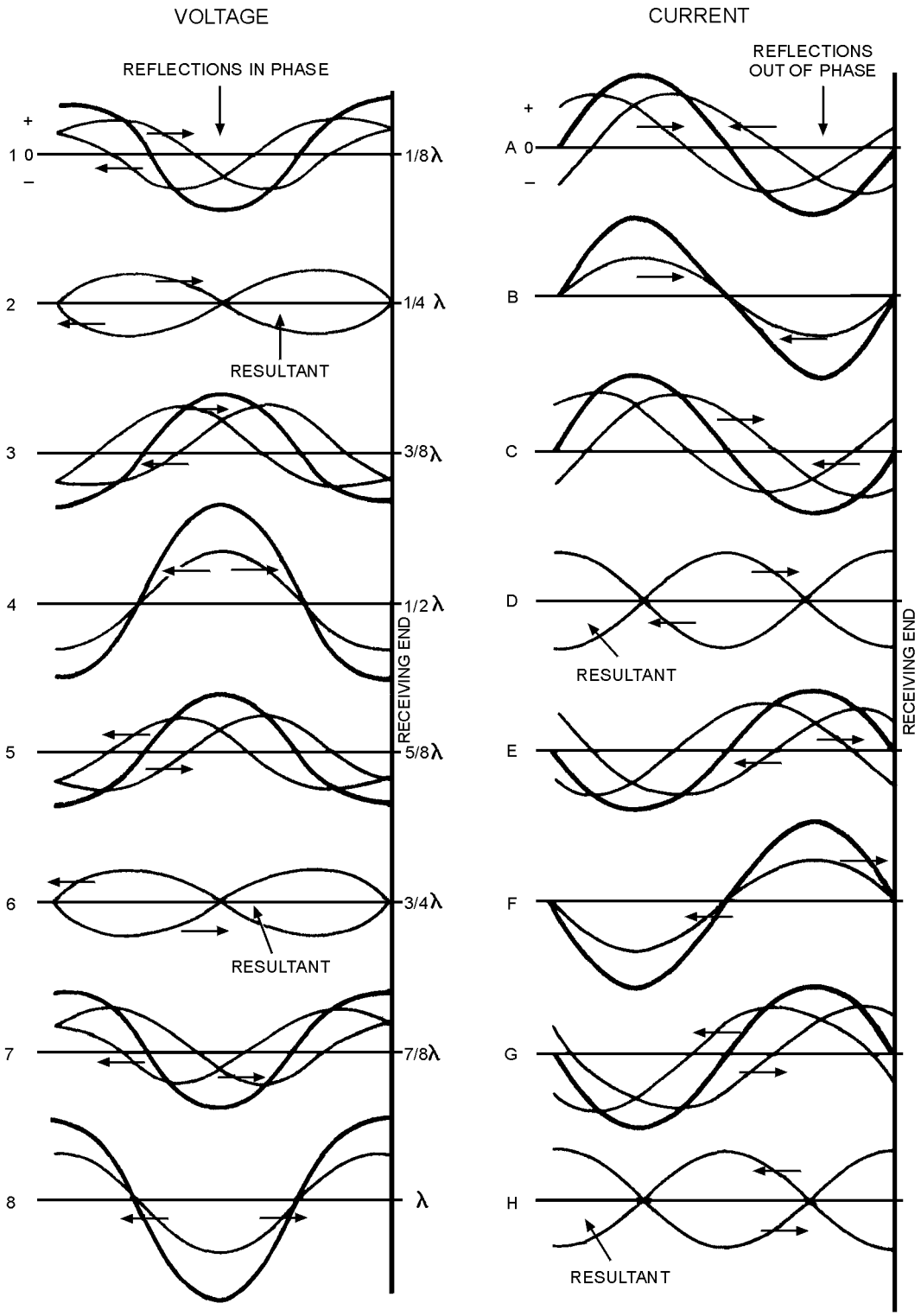


Figure 3-27.—Instantaneous values of incident and reflected waves on an open-ended line.

In waveform (4), the incident wave is at a maximum negative value at the end of the line. The wave has moved another 45 degrees to the right from the wave in the preceding illustration. The reflected wave has also moved 45 degrees, but to the left. The reflected wave is in phase with the incident wave. The resultant of these two waves, shown by the dark line, again has a negative maximum at the end of the line and a positive maximum 1/2λ from the end of the line. Notice that these maxima have a greater amplitude than those in waveform (3).

In waveform (5), the incident wave has moved another 45 degrees to the right and the reflected wave 45 degrees to the left. The resultant again is maximum negative at the end and positive maximum 1/2λ from the end. The maxima are lower than those in waveform (4). In waveform (6), the incident and reflected wave have moved another 1/8λ. The two waves again are 180 degrees out of phase, giving a resultant wave with no amplitude. The incident and reflected waves continue moving in opposite directions, adding to produce the resultant waveshapes shown in waveforms (7) and (8). Notice that the maximum voltage in each resultant wave is at the end and 1/2λ from the end.

Study each part of figure 3-27 carefully and you will get a clear picture of how the resultant waveforms of voltage are produced. You will also see that the resultant voltage wave on an open-ended line is always zero at 1/4λ and 3/4λ from the end of the transmission line. Since the zero and maximum points are always in the same place, the resultant of the incident and the reflected wave is called a STANDING WAVE of voltage.

The right-hand column in figure 3-27 shows the *current* waveshapes on the open-ended line. Since the current is reflected out of phase at an open end, the resultant waveshapes differ from those for voltage. The two out-of-phase components always cancel at the end of the transmission line, so the resultant is always zero at that point. If you check all the resultant waveshapes shown in the right-hand column of figure 3-27, you will see that a zero point always occurs at the end and at a point 1/2λ from the end. Maximum voltages occur 1/4λ and 3/4λ from the end.

When an ac meter is used to measure the voltages and currents along a line, the polarity is not indicated. If you plot all the current and voltage readings along the length of the line, you will get curves like the ones shown in figure 3-28. Notice that all are positive. These curves are the conventional method of showing current and voltage standing waves on rf lines.

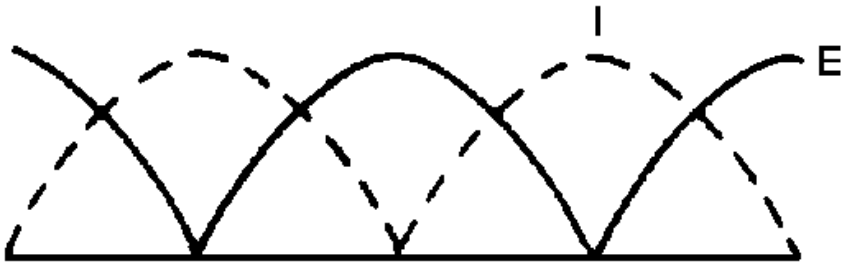


Figure 3-28.—Conventional picture of standing waves.

When an rf line is terminated in a short circuit, reflection is complete, but the effect on voltage and current differs from that in an open-ended line. Voltage is reflected in opposite phase, while current is reflected in phase. Again refer to the series of pictures shown in figure 3-27. However, this time the left column represents *current*, since it shows reflection in phase; and the right column of pictures now represents the *voltage* changes on the shorted line, since it shows reflection out of phase.

The composite diagram in figure 3-29 shows all resultant curves on a full-wavelength section of line over a complete cycle. Notice that the amplitude of the voltage varies between zero and maximum in both directions at the center and at both ends as well but, one-fourth of the distance from each end the voltage is always zero. The resultant waveshape is referred to as a standing wave of voltage. Standing waves, then, are caused by reflections, which occur only when the line is not terminated in its characteristic impedance.

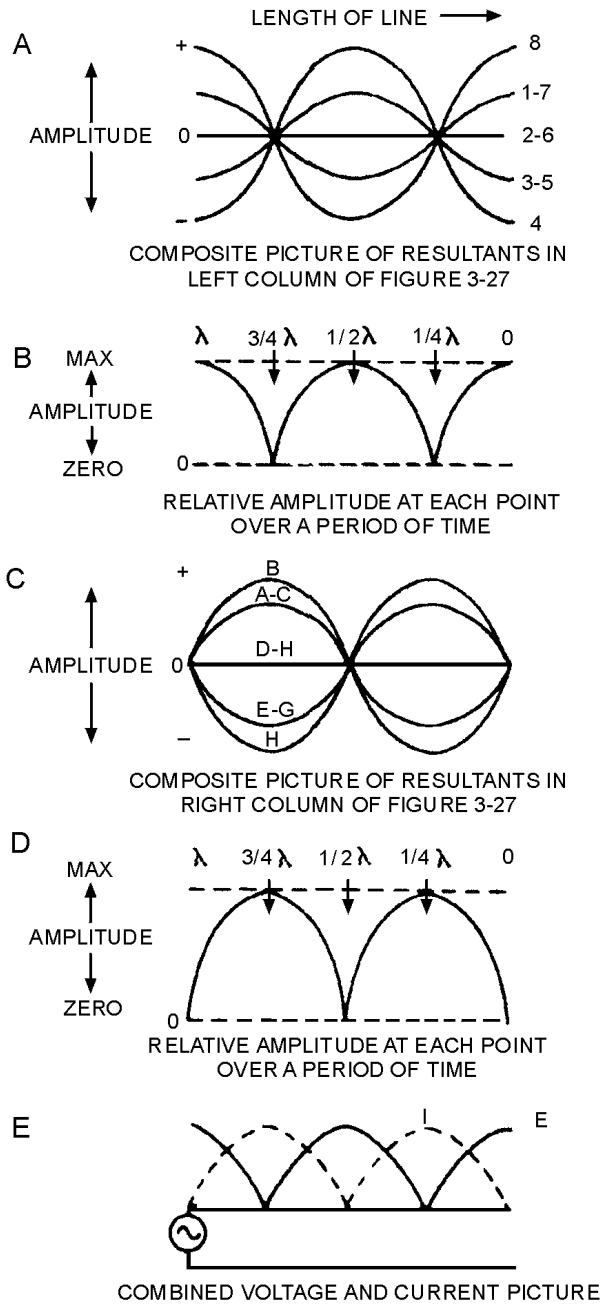


Figure 3-29.—Composite results of instantaneous waves.

The voltage at the center and the ends varies at a sinusoidal rate between the limits shown. At the one-fourth the three-fourths points, the voltage is always zero. A continuous series of diagrams such as these is difficult to see with conventional test equipment, which reads the effective or average voltage over several cycles. The curve of amplitude over the length of line for several cycles is shown in figure 3-29, view B. A meter will read zero at the points shown and will show a maximum voltage at the center, no matter how many cycles pass.

As shown in view D, the amplitude varies along the length of the line. In this case it is zero at the end and center but maximum at the one-fourth and three-fourths points. The entire diagram of the open-ended line conditions is shown in view E. The standing waves of voltage and current appear together. Observe that one is maximum when the other is minimum. The current and voltage standing waves are one-quarter cycle, or 90 degrees, out of phase with one another.

REFLECTION OF AC VOLTAGE FROM A SHORT CIRCUIT

Reflection is complete when an rf line is terminated in a short circuit, but the effect on voltage and current differs from the effect obtained in an open-ended line. Voltage is reflected in opposite phase, while current is reflected in phase. Again look at the series of diagrams in figure 3-27. The left column represents current, and the right column shows voltage changes on the shorted line. The standard representation of standing waves on a shorted line is shown in figure 3-30; the voltage is a solid line, and the current is a dashed line. The voltage is zero at the end and center (1/2λ) and maximum at the 1/4λ and 3/4λ points, while the current is maximum at the end and center and minimum at the 1/4λ and 3/4λ points.

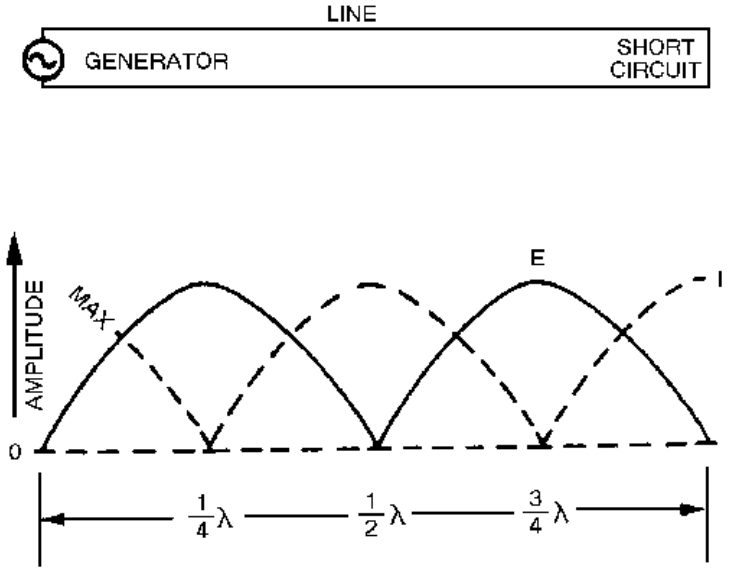


Figure 3-30.—Standing waves on a shorted line.

As we discussed voltage and current waves on transmission lines, we pointed out several differences between open and shorted lines. Basic differences also appear in the standing-wave patterns for open and shorted lines. You can see these differences by comparing figure 3-29, view E, and figure 3-30. Notice that the current and voltage standing waves are shifted 90 degrees with respect to the termination. At the open end of a line, voltage is maximum (zero if there are no losses in the line). At a short circuit, current is maximum and voltage is minimum.

Q23. Two types of waves are formed on a transmission line. What names are given to these waves?

Q24. In figure 3-27, which waveforms on the left have a resultant wave of zero, and what is indicated by these waves?

Q25. On an open-ended transmission line, the voltage is always zero at what distance from each end of the line?

TERMINATING A TRANSMISSION LINE

A transmission line is either NONRESONANT or RESONANT. First, let us define the terms nonresonant lines and resonant lines. A nonresonant line is a line that has no standing waves of current and voltage. A resonant line is a line that has standing waves of current and voltage.

Nonresonant Lines

A nonresonant line is either infinitely long or terminated in its characteristic impedance. Since no reflections occur, all the energy traveling down the line is absorbed by the load which terminates the line. Since no standing waves are present, this type of line is sometimes spoken of as a FLAT line. In addition, because the load impedance of such a line is equal to Z_0, no special tuning devices are required to effect a maximum power transfer; hence, the line is also called an UNTUNED line.

Resonant Lines

A resonant line has a finite length and is not terminated in its characteristic impedance. Therefore reflections of energy do occur. The load impedance is different from the Z_0 of the line; therefore, the input impedance may not be purely resistive but may have reactive components. Tuning devices are used to eliminate the reactance and to bring about maximum power transfer from the source to the line. Therefore, a resonant line is sometimes called a TUNED line. The line also may be used for a resonant or tuned circuit.

A resonant line is sometimes said to be resonant at an applied frequency. This means that at one frequency the line acts as a resonant circuit. It may act either as a high-resistive circuit (parallel resonant) or as a low-resistive circuit (series resonant). The line may be made to act in this manner by either open- or short-circuiting it at the output end and cutting it to some multiple of a quarter-wavelength.

At the points of voltage maxima and minima on a short-circuited or open-circuited line, the line impedance is resistive. On a short-circuited line, each point at an odd number of quarter-wavelengths from the receiving end has a high impedance (figure 3-31, view A). If the frequency of the applied voltage to the line is varied, this impedance decreases as the effective length of the line changes. This variation is exactly the same as the change in the impedance of a parallel-resonant circuit when the applied frequency is varied.

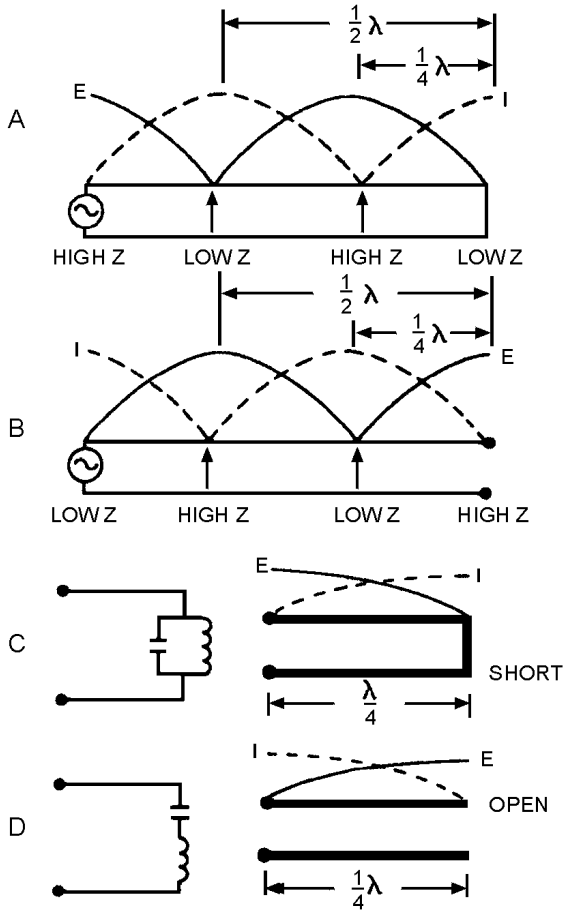


Figure 3-31.—Sending-end impedance of various lengths and terminations.

At all even numbered quarter-wavelength points from the short circuit, the impedance is extremely low. When the frequency of the voltage applied to the line is varied, the impedance at these points increases just as the impedance of a series-resonant circuit varies when the frequency applied to it is changed. The same is true for an open-ended line (figure 3-31, view B) except that the points of high and low impedance are reversed.

At this point let us review some of the characteristics of resonant circuits so we can see how resonant line sections may be used in place of LC circuits.

A PARALLEL-RESONANT circuit has the following characteristics:

- At resonance the impedance appears as a very high resistance. A loss-free circuit has infinite impedance (an open circuit). Other than at resonance, the impedance decreases rapidly.

- If the circuit is resonant at a point above the generator frequency (the generator frequency is too low), more current flows through the coil than through the capacitor. This happens because X_L decreases with a decrease in frequency but X_C increases.

A SERIES-RESONANT circuit has these characteristics:

- At resonance the impedance appears as a very low resistance. A loss-free circuit has zero impedance (a short circuit). Other than at resonance the impedance increases rapidly.

- If the circuit is resonant at a point above the generator frequency (the generator frequency is too low), then X_C is larger than X_L and the circuit acts capacitively.

- If the circuit is resonant at a point below the generator frequency (the generator frequency is too high), then X_L is larger than X_C and the circuit acts inductively.

Since the impedance a generator sees at the quarter-wave point in a shorted line is that of a parallel-resonant circuit, a shorted quarter-wave- length of line may be used as a parallel-resonant circuit (figure 3-31, view C). An open quarter-wavelength of line may be used as a series-resonant circuit (view D). The Q of such a resonant line is much greater than can be obtained with lumped capacitance and inductance.

Impedance for Various Lengths of Open Lines

In figure 3-32, the impedance (Z) the generator sees for various lengths of line is shown at the top. The curves above the letters of various heights show the relative value of the impedances presented to the generator for the various line lengths. The circuit symbols indicate the equivalent electrical circuits for the transmission lines at each particular length. The standing waves of voltage and current are shown on each length of line.

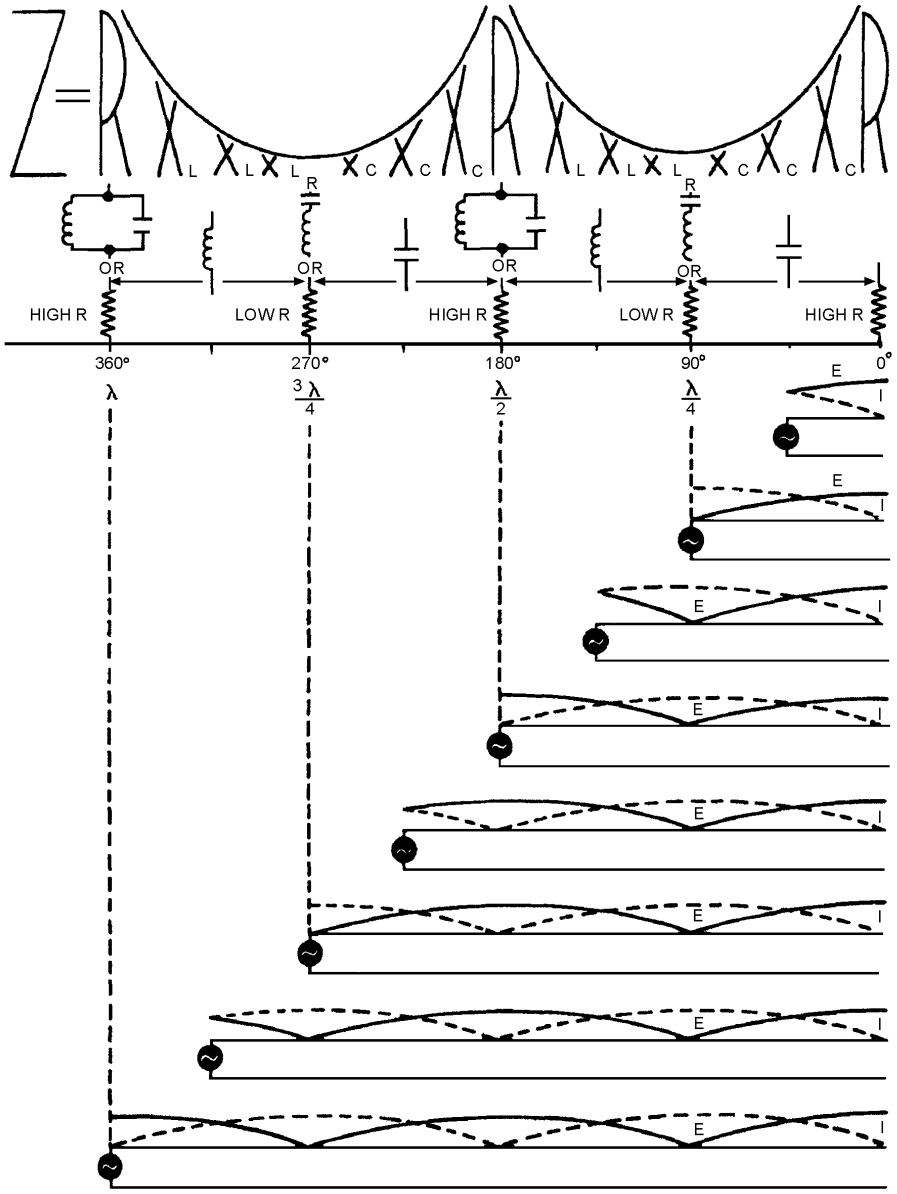


Figure 3-32.—Voltage, current, and impedance on open line.

At all odd quarter-wave points (1/4λ, 3/4λ, etc.), the voltage is minimum, the current is maximum, and the impedance is minimum. Thus, at all odd quarter-wave points, the open-ended transmission line acts as a series-resonant circuit. The impedance is equivalent to a very low resistance, prevented from being zero only by small circuit losses.

At all even quarter-wave points (1/2λ, 1λ, 3/2λ, etc.), the voltage is maximum, the current is minimum, and the impedance is maximum. Comparison of the line with an LC resonant circuit shows that at an even number of quarter-wavelengths, an open line acts as a parallel-resonant circuit. The impedance is therefore an extremely high resistance.

In addition, resonant open lines may also act as nearly pure capacitances or inductances. The illustration shows that an open line less than a quarter-wavelength long acts as a capacitance. Also, it acts

as an inductance from 1/4 to 1/2 wavelength, as a capacitance from 1/2 to 3/4 wavelength, and as an inductance from 3/4 to 1 wavelength, etc. A number of open transmission lines, with their equivalent circuits, are shown in the illustration.

Impedance of Various Lengths of Shorted Lines

Follow figure 3-33 as we study the shorted line. At the odd quarter-wavelength points, the voltage is high, the current is low, and the impedance is high. Since these conditions are similar to those found in a parallel-resonant circuit, the shorted transmission line acts as a parallel-resonant circuit at these lengths.

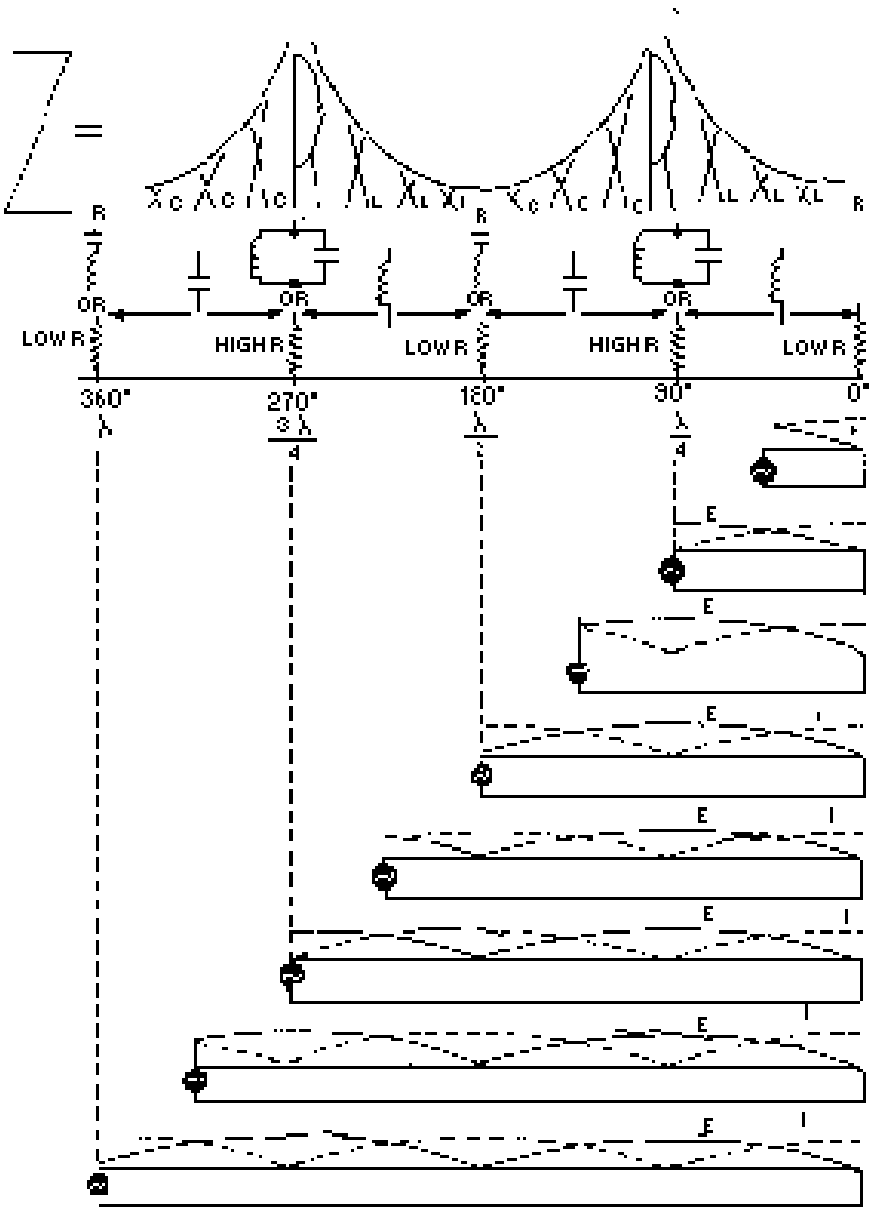


Figure 3-33.—Voltage, current, and impedance on shorted line.

At the even quarter-wave points voltage is minimum, current is maximum, and impedance is minimum. Since these characteristics are similar to those of a series-resonant LC circuit, a shorted transmission line whose length is an even number of quarter-wavelengths acts as a series-resonant circuit.

Resonant shorted lines, like open-end lines, also may act as pure capacitances or inductances. The illustration shows that a shorted line less than 1/4 wavelength long acts as an inductance. A shorted line with a length of from 1/4 to 1/2 wavelength acts as a capacitance. From 1/2 to 3/4 wavelength, the line acts as an inductance; and from 3/4 to 1 wavelength, it acts as a capacitance, and so on. The equivalent circuits of shorted lines of various lengths are shown in the illustration. Thus, properly chosen line segments may be used as parallel-resonant, series-resonant, inductive, or capacitive circuits.

STANDING WAVES ON A TRANSMISSION LINE

There is a large variety of terminations for rf lines. Each type of termination has a characteristic effect on the standing waves on the line. From the nature of the standing waves, you can determine the type of termination that produces the waves.

TERMINATION IN Z_0

Termination in Z_0 (characteristic impedance) will cause a constant reading on an ac meter when it is moved along the length of the line. As illustrated in figure 3-34, view A, the curve, provided there are no losses in the line, will be a straight line. If there are losses in the line, the amplitude of the voltage and current will diminish as they move down the line (view B). The losses are due to dc resistance in the line itself.

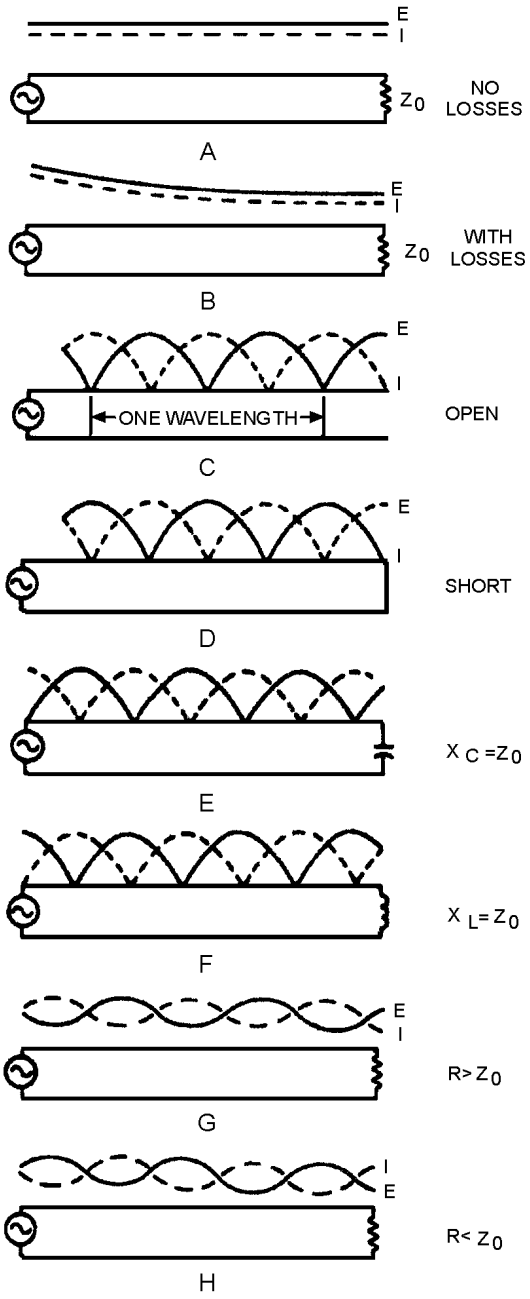


Figure 3-34.—Effects of various terminations on standing waves.

TERMINATION IN AN OPEN CIRCUIT

In an open-circuited rf line (figure 3-34, view C), the voltage is maximum at the end, but the current is minimum. The distance between two adjacent zero current points is 1/2λ, and the distance between alternate zero current points is 1λ. The voltage is zero at a distance of 1/4λ from the end of the line. This is true at any frequency. A voltage peak occurs at the end of the line, at 1/2λ from the end, and at each 1/2λ thereafter.

TERMINATION IN A SHORT CIRCUIT

On the line terminated in a short circuit, shown in figure 3-34, view D, the voltage is zero at the end and maximum at 1/4λ from the end. The current is maximum at the end, zero at 1/4λ from the end, and alternately maximum and zero every 1/4λ thereafter.

TERMINATION IN CAPACITANCE

When a line is terminated in capacitance, the capacitor does not absorb energy, but returns all of the energy to the circuit. This means there is 100 percent reflection. The current and voltage relationships are somewhat more involved than in previous types of termination. For this explanation, assume that the capacitive reactance is equal to the Z_0 of the line. Current and voltage are in phase when they arrive at the end of the line, but in flowing through the capacitor and the characteristic impedance (Z_0) connected in series, they shift in phase relationship. Current and voltage arrive in phase and leave out of phase. This results in the standing-wave configuration shown in figure 3-34, view E. The standing wave of voltage is minimum at a distance of exactly 1/8λ from the end. If the capacitive reactance is greater than Z_0 (smaller capacitance), the termination looks more like an open circuit; the voltage minimum moves away from the end. If the capacitive reactance is smaller than Z_0, the minimum moves toward the end.

TERMINATION IN INDUCTANCE

When the line is terminated in an inductance, both the current and voltage shift in phase as they arrive at the end of the line. When X_L is equal to Z_0, the resulting standing waves are as shown in figure 3-34, view F. The current minimum is located 1/8λ from the end of the line. When the inductive reactance is increased, the standing waves appear closer to the end. When the inductive reactance is decreased, the standing waves move away from the end of the line.

TERMINATION IN A RESISTANCE NOT EQUAL TO THE CHARACTERISTIC IMPEDANCE (Z_0)

Whenever the termination is not equal to Z_0, reflections occur on the line. For example, if the terminating element contains resistance, it absorbs some energy, but if the resistive element does not equal the Z_0 of the line, some of the energy is reflected. The amount of voltage reflected may be found by using the equation:

$$E_R = E_i\left(\frac{R_L - Z_0}{R_L + Z_0}\right)$$

Where:

E_R = the reflected voltage

E_i = the incident voltage

R_R = the terminating resistance

Z_0= the characteristic impedance of the line

If you try different values of R_L in the preceding equation, you will find that the reflected voltage is equal to the incident voltage only when R_L equals 0 or is infinitely large. When R_L equals Z_0, no reflected voltage occurs. When R_L is greater than Z_0, E_R is positive, but less than E_i. As R_L increases and

approaches an infinite value, E_R increases and approaches E_i in value. When R_L is smaller than Z_0, E_R has a negative value. This means that the reflected voltage is of opposite polarity to the incident wave at the termination of the line. As R_L approaches zero, E_R approaches E_i in value. The smaller the value of E_R, the smaller is the peak amplitude of the standing waves and the higher are the minimum values.

TERMINATION IN A RESISTANCE GREATER THAN Z_0

When R_L is greater than Z_0, the end of the line is somewhat like an open circuit; that is, standing waves appear on the line. The voltage maximum appears at the end of the line and also at half-wave intervals back from the end. The current is minimum (not zero) at the end of the line and maximum at the odd quarter-wave points. Since part of the power in the incident wave is consumed by the load resistance, the minimum voltage and current are less than for the standing waves on an open-ended line. Figure 3-34, view G, illustrates the standing waves for this condition.

TERMINATION IN A RESISTANCE LESS THAN Z_0

When R_L is less than Z_0, the termination appears as a short circuit. The standing waves are shown in figure 3-34, view H. Notice that the line terminates in a current LOOP (peak) and a voltage NODE (minimum). The values of the maximum and minimum voltage and current approach those for a shorted line as the value of R_L approaches zero.

A line does not have to be any particular length to produce standing waves; however, it cannot be an infinite line. Voltage and current must be reflected to produce standing waves. For reflection to occur, a line must not be terminated in its characteristic impedance. Reflection occurs on lines terminated in opens, shorts, capacitances, and inductances, because no energy is absorbed by the load. If the line is terminated in a resistance not equal to the characteristic impedance of the line, some energy will be absorbed and the rest will be reflected.

The voltage and current relationships for open-ended and shorted lines are opposite to each other, as shown in figure 3-34, views C and D. The points of maximum and minimum voltage and current are determined from the output end of the line, because reflection always begins at that end.

Q26. A nonresonant line is a line that has no standing waves of current and voltage on it and is considered to be flat. Why is this true?

Q27. On an open line, the voltage and impedance are maximum at what points on the line?

STANDING-WAVE RATIO

The measurement of standing waves on a transmission line yields information about equipment operating conditions. Maximum power is absorbed by the load when $Z_L = Z_0$. If a line has no standing waves, the termination for that line is correct and maximum power transfer takes place.

You have probably noticed that the variation of standing waves shows how near the rf line is to being terminated in Z_0. A wide variation in voltage along the length means a termination far from Z_0. A small variation means termination near Z_0. Therefore, the ratio of the maximum to the minimum is a measure of the perfection of the termination of a line. This ratio is called the STANDING-WAVE RATIO (swr) and is always expressed in whole numbers. For example, a ratio of 1:1 describes a line terminated in its characteristic impedance (Z_0).

Voltage Standing-Wave Ratio

The ratio of maximum voltage to minimum voltage on a line is called the VOLTAGE STANDING-WAVE RATIO (vswr). Therefore:

$$\text{vswr} = \frac{E_{max}}{E_{min}}$$

The vertical lines in the formula indicate that the enclosed quantities are absolute and that the two values are taken without regard to polarity. Depending on the nature of the standing waves, the numerical value of vswr ranges from a value of 1 ($Z_L = Z_0$, no standing waves) to an infinite value for theoretically complete reflection. Since there is always a small loss on a line, the minimum voltage is never zero and the vswr is always some finite value. However, if the vswr is to be a useful quantity, the power losses along the line must be small in comparison to the transmitted power.

Power Standing-Wave Ratio

The square of the voltage standing-wave ratio is called the POWER STANDING-WAVE RATIO (pswr). Therefore:

$$\text{pswr} = \frac{P_{max}}{P_{min}}$$

This ratio is useful because the instruments used to detect standing waves react to the square of the voltage. Since power is proportional to the square of the voltage, the ratio of the square of the maximum and minimum voltages is called the power standing-wave ratio. In a sense, the name is misleading because the power along a transmission line does not vary.

Current Standing-Wave Ratio

The ratio of maximum to minimum current along a transmission line is called CURRENT STANDING-WAVE RATIO (iswr). Therefore:

$$\text{iswr} = \frac{I_{max}}{I_{min}}$$

This ratio is the same as that for voltages. It can be used where measurements are made with loops that sample the magnetic field along a line. It gives the same results as vswr measurements.

Q28. At what point on an open-circuited rf line do voltage peaks occur?

Q29. What is the square of the voltage standing-wave ratio called?

Q30. What does vswr measure?

SUMMARY

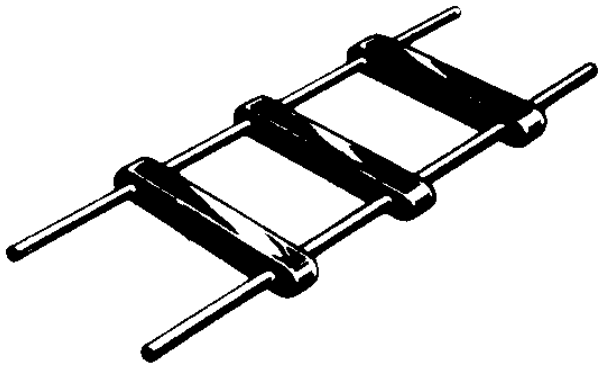
This chapter has presented information on the characteristics of transmission lines. The information that follows summarizes the important points of this chapter.

TRANSMISSION LINES are devices for guiding electrical energy from one point to another.

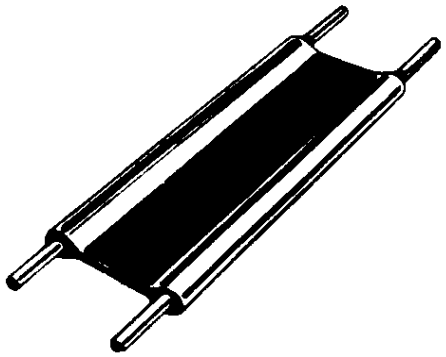
INPUT IMPEDANCE is the ratio of voltage to current at the input end of a transmission line.

OUTPUT IMPEDANCE is the ratio of voltage to current at the output end of the line.

TWO-WIRE OPEN LINES are parallel lines and have uses such as power lines, rural telephone lines, and telegraph lines. This type of line has high radiation losses and is subject to noise pickup.

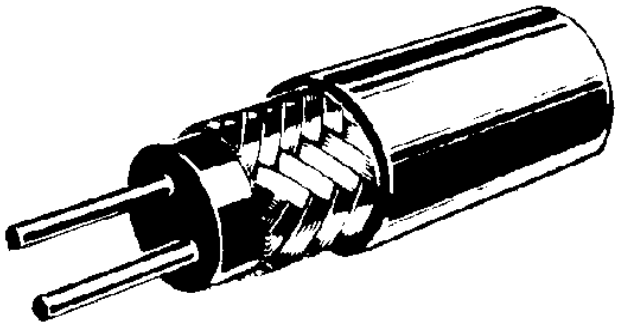


TWIN LEAD has parallel lines and is most often used to connect televisions to their antennas.

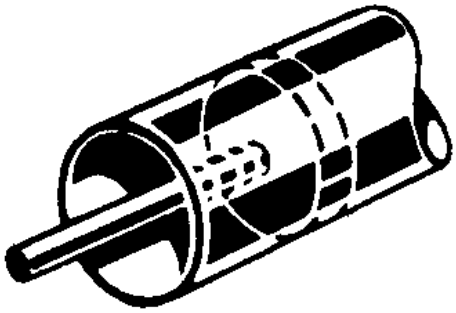


A **TWISTED PAIR** consists of two insulated wires twisted together. This line has high insulation loss.

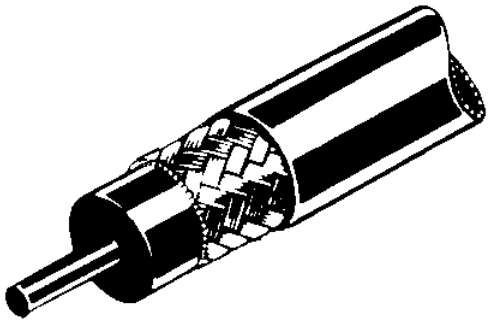
A **SHIELDED PAIR** has parallel conductors separated by a solid dielectric and surrounded by copper braided tubing. The conductors are balanced to ground.



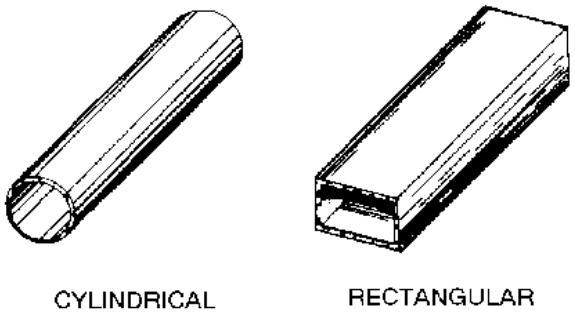
RIGID COAXIAL LINE contains two concentric conductors insulated from each other by spacers. Some rigid coaxial lines are pressurized with an inert gas to prevent moisture from entering. High-frequency losses are less than with other lines.



FLEXIBLE COAXIAL LINES consist of a flexible inner conductor and a concentric outer conductor of metal braid. The two are separated by a continuous insulating material.



WAVEGUIDES are hollow metal tubes used to transfer energy from one point to another. The energy travels slower in a waveguide than in free space.



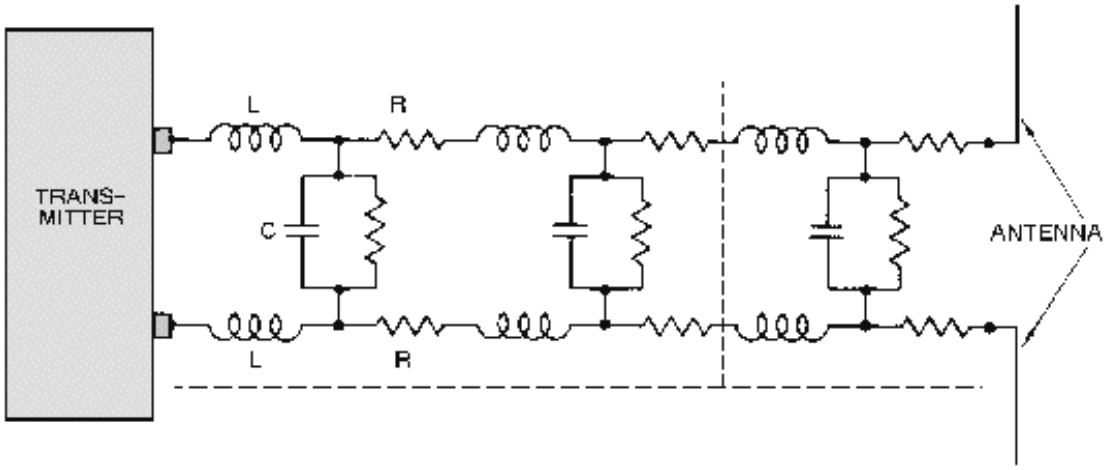
COPPER LOSSES can result from power (I^2R) loss, in the form of heat, or skin effect. These losses decrease the conductivity of a line.

DIELECTRIC LOSSES are caused by the heating of the dielectric material between conductors, taking power from the source.

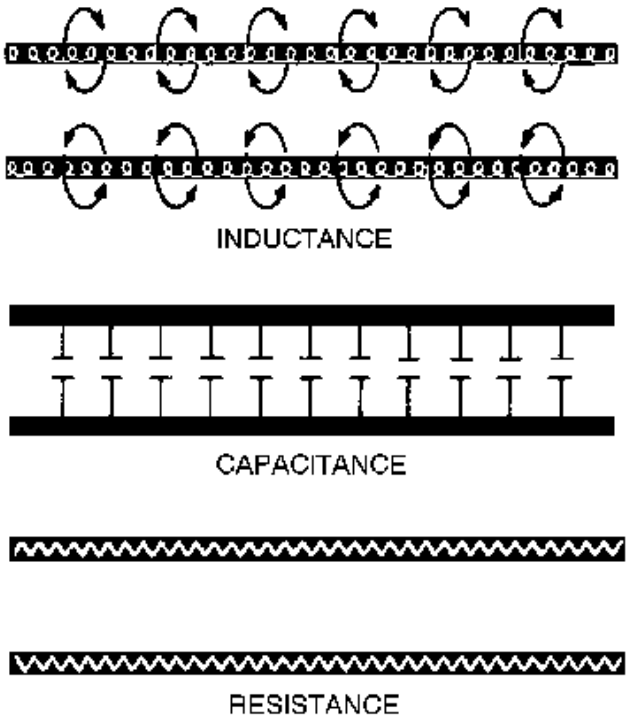
RADIATION and **INDUCTION LOSSES** are caused by part of the electromagnetic fields of a conductor being dissipated into space or nearby objects.

A transmission line is either electrically **LONG** or **SHORT** if its physical length is not equal to $1/4\lambda$ for the frequency it is to carry.

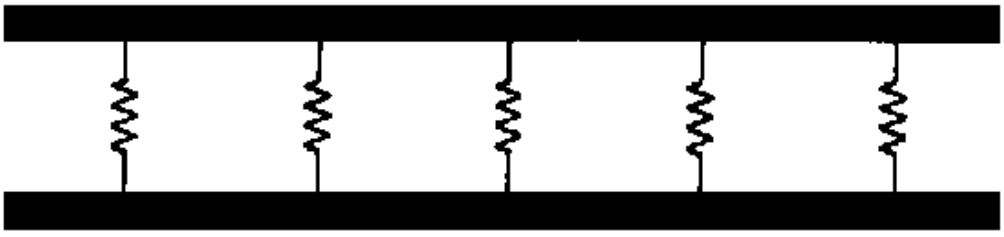
LUMPED CONSTANTS are theoretical properties (inductance, resistance, and capacitance) of a transmission line that are lumped into a single component.



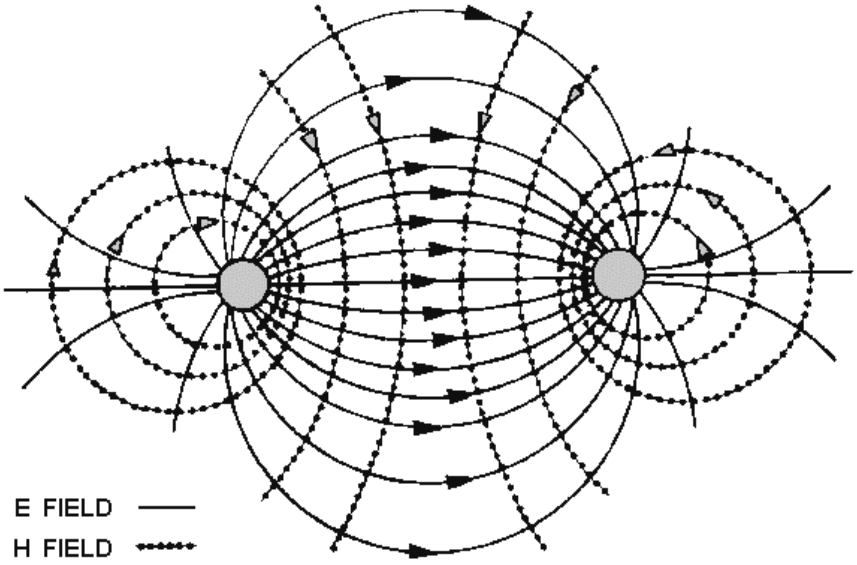
DISTRIBUTED CONSTANTS are constants of inductance, capacitance and resistance that are distributed along the transmission line.



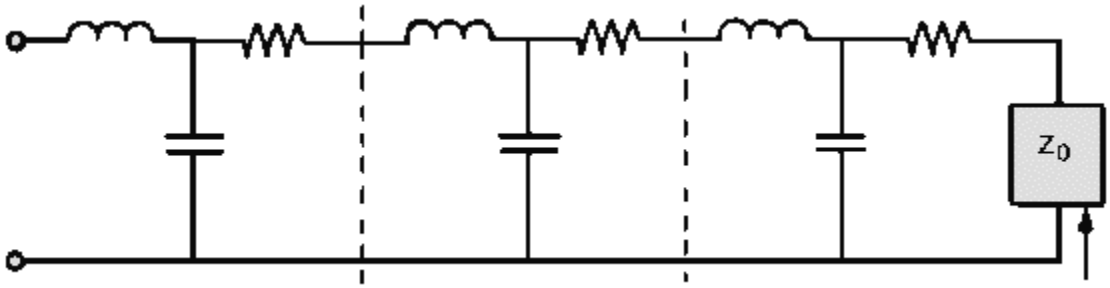
LEAKAGE CURRENT flows between the wires of a transmission line through the dielectric. The dielectric acts as a resistor.



An **ELECTROMAGNETIC FIELD** exists along transmission line when current flows through it.



CHARACTERISTIC IMPEDANCE, Z_0, is the ratio of E to I at every point along the line. For maximum transfer of electrical power, the characteristic impedance and load impedance must be matched.



The **VELOCITY** at which a wave travels over a given length of transmission line can be found by using the formula:

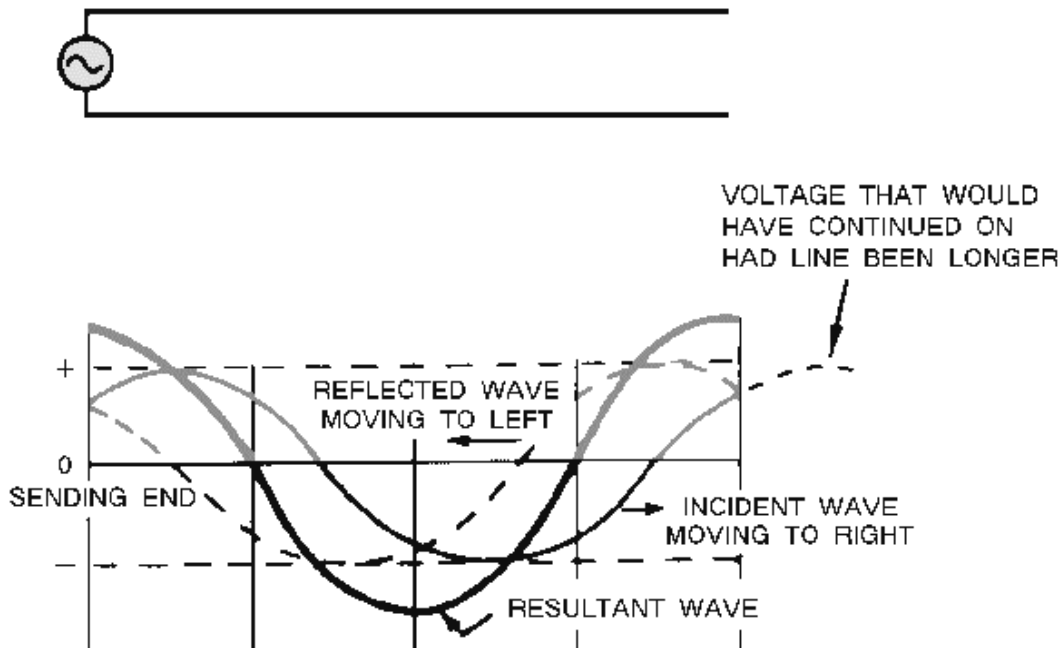
$$V = \frac{D}{\sqrt{LC}}$$

A transmission line that is not terminated in its characteristic impedance is said to be **FINITE**.

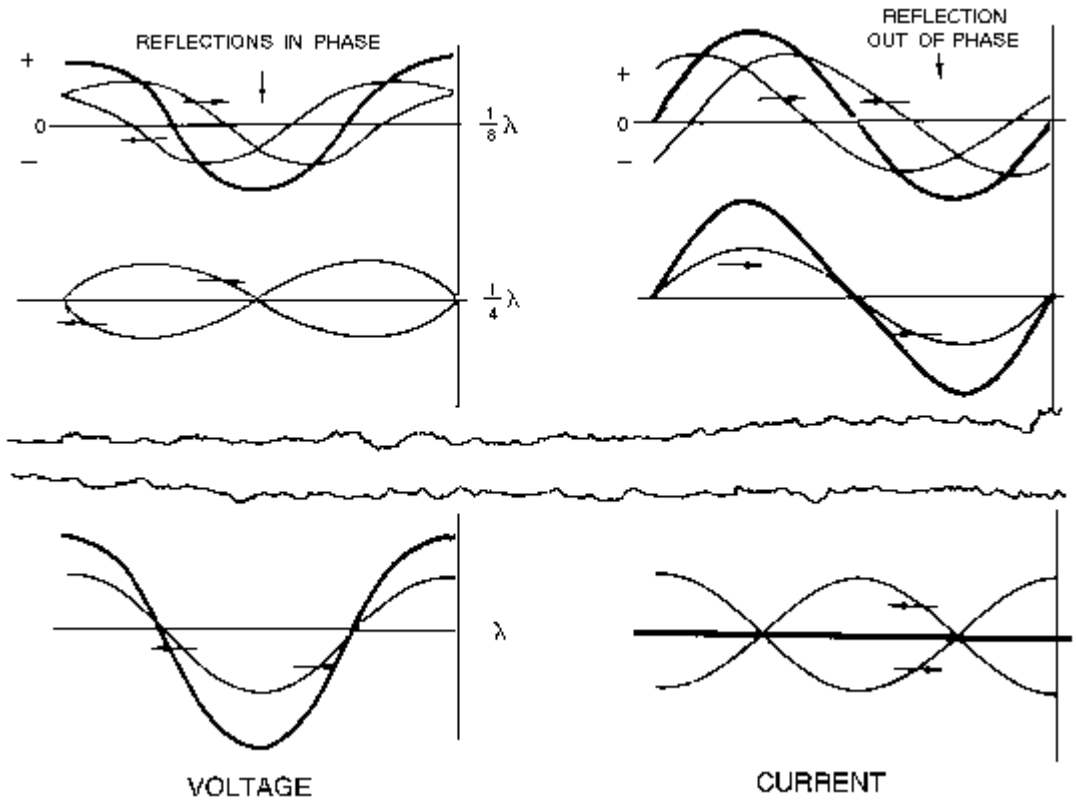
When dc is applied to an **OPEN-ENDED** line, the voltage is reflected back from the open end without any change in polarity, amplitude, or shape. Current is reflected back with the same amplitude and shape but with opposite polarity.

When dc is applied to a **SHORT-CIRCUITED** line, the current is reflected back with the same amplitude, and polarity. The voltage is reflected back with the same amplitude but with opposite polarity.

When ac is applied to an **OPEN-END** line, voltage is always reflected back in phase with the incident wave and current is reflected back out of phase.



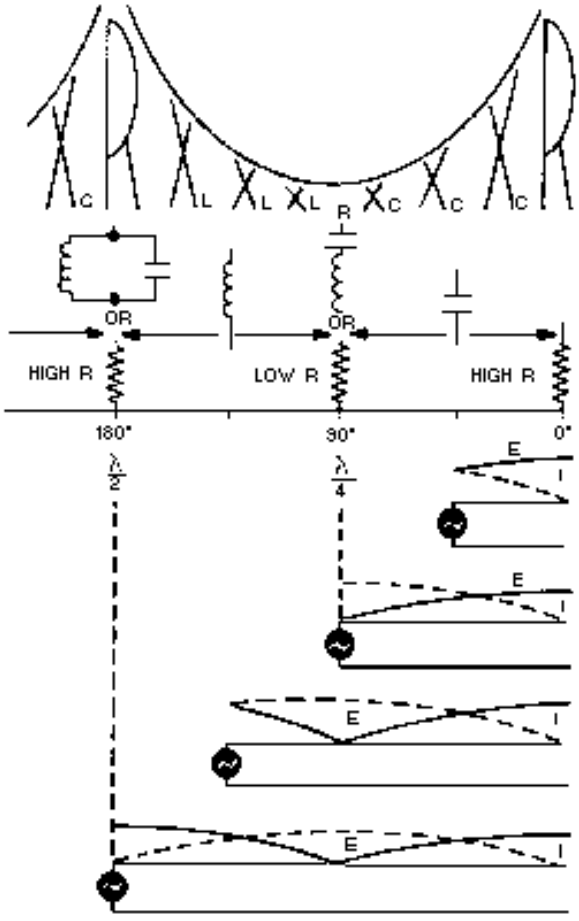
When ac is applied to a **SHORT-CIRCUITED** line, voltage is reflected in opposite phase, while current is reflected in phase.



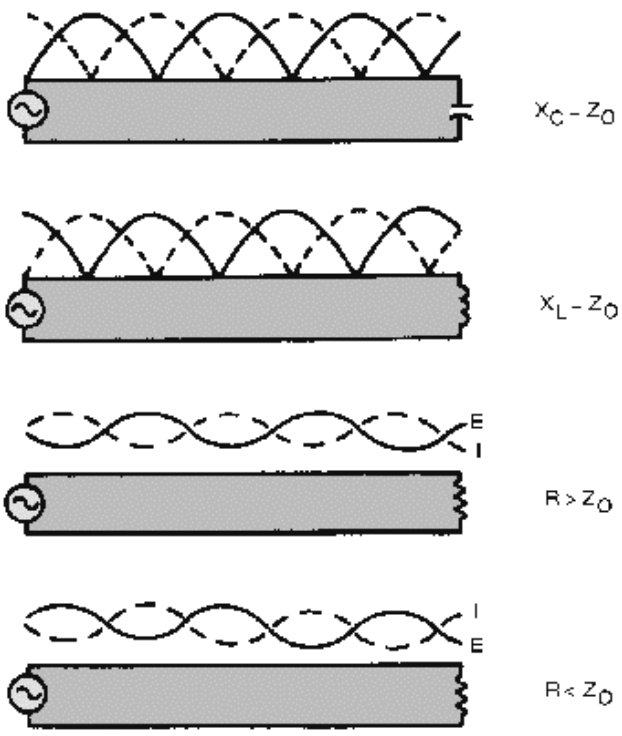
A **NONRESONANT** line has **NO STANDING WAVES** of current and voltage and is either infinitely long or terminated in its characteristic impedance.

A **RESONANT** line has **STANDING WAVES** of current and voltage and is of finite length and is NOT terminated in its characteristic impedance.

On an open-ended resonant line, and at all odd 1/4λ points, the voltage is minimum, the current is maximum, and the impedance is minimum. At all even 1/4λ points, the voltage is maximum, the current is minimum and the impedance is maximum.



There are a variety of **TERMINATIONS** for rf lines. Each termination has an effect on the standing waves on the line.



A transmission line can be terminated in its characteristic impedance as an open- or short-circuit, or in capacitance or inductance.

Whenever the termination on a transmission line is NOT EQUAL TO Z_0, there are reflections on the line. The amount of voltage reflected may be found by using the equation:

$$E_R = E_i\left(\frac{R_L - Z_0}{R_L + Z_0}\right)$$

When the termination on a transmission line EQUALS Z_0, there is NO reflected voltage.

The measurement of standing waves on a transmission line yields information about operating conditions. If there are NO standing waves, the termination for that line is correct and maximum power transfer takes place.

The **STANDING WAVE RATIO** is the measurement of maximum voltage (current) to minimum voltage (current) on a transmission line and measures the perfection of the termination of the line. A ratio of 1:1 describes a line terminated in its characteristic impedance.

ANSWERS TO QUESTIONS Q1. THROUGH Q30.

A1. Transmission line.

A2. Input end, generator end, transmitter end, sending end, and source.

A3. Output end, receiving end, load end and sink.

A4. Parallel two-wire, twisted pair, shielded pair, coaxial line and waveguide.

A5. Power lines, rural telephone lines, and telegraph lines.

A6. High radiation losses and noise pickup.

A7. Twin lead.

A8. The conductors are balanced to ground.

A9. Air coaxial (rigid) and solid coaxial (flexible).

A10. The ability to minimize radiation losses.

A11. Expensive to construct, must be kept dry, and high frequency losses limit the practical length of the line.

A12. Cylindrical and rectangular.

A13. Copper, dielectric, and radiation.

A14. Copper losses.

A15. Dielectric losses.

A16. λ = 20 meters.

A17. (1) Type of line used, (2) dielectric in the line, and (3) length of line.

A18. Inductance is expressed in microhenrys per unit length, capacitance is expressed in picofarads per unit length, and resistance is expressed in ohms per unit length.

A19. The small amount of current that flows through the dielectric between two wires of a transmission line and is expressed in micromhos per unit length.

A20. When the characteristic impedance of the transmission line and the load impedance are equal.

A21. Z_0 and it is the ratio of E to I at every point along the line.

A22. Between 50 and 600 ohms.

A23. Incident waves from generator to load. Reflected waves from load back to generator.

A24. 2 and 6 have zero resultant wave and they indicate that the incident and reflected waves are 180 degrees out of phase at all parts.

A25. One-fourth the distance from each end of the line.

A26. The load impedance of such a line is equal to Z_0.

A27. Even quarter-wave points (1/2λ, 1λ, 3/2λ, etc.).

A28. At 1/2 wavelength from the end and at every 1/2 wavelength along the line.

A29. Power standing-wave ratio (pswr).

A30. The existence of voltage variations on a line.

CHAPTER 4

ANTENNAS

LEARNING OBJECTIVES

Upon completion of this chapter you will be able to:

1. State the basic principles of antenna radiation and list the parts of an antenna.
2. Explain current and voltage distribution on an antenna.
3. Describe how electromagnetic energy is radiated from an antenna.
4. Explain polarization, gain, and radiation resistance characteristics of an antenna.
5. Describe the theory of operation of half- wave and quarter-wave antennas.
6. List the various array antennas.
7. Describe the directional array antennas presented and explain the basic operation of each.
8. Identify various special antennas presented, such as long-wire, V, rhombic, turnstile, ground-plane, and corner-reflector; describe the operation of each.
9. List safety precautions when working aloft and around antennas.

INTRODUCTION

If you had been around in the early days of electronics, you would have considered an ANTENNA (AERIAL) to be little more than a piece of wire strung between two trees or upright poles. In those days, technicians assumed that longer antennas automatically provided better reception than shorter antennas. They also believed that a mysterious MEDIUM filled all space, and that an antenna used this medium to send and receive its energy. These two assumptions have since been discarded. Modern antennas have evolved to the point that highly directional, specially designed antennas are used to relay worldwide communications in space through the use of satellites and Earth station antennas (fig. 4-1). Present transmission theories are based on the assumption that space itself is the only medium necessary to propagate (transmit) radio energy.

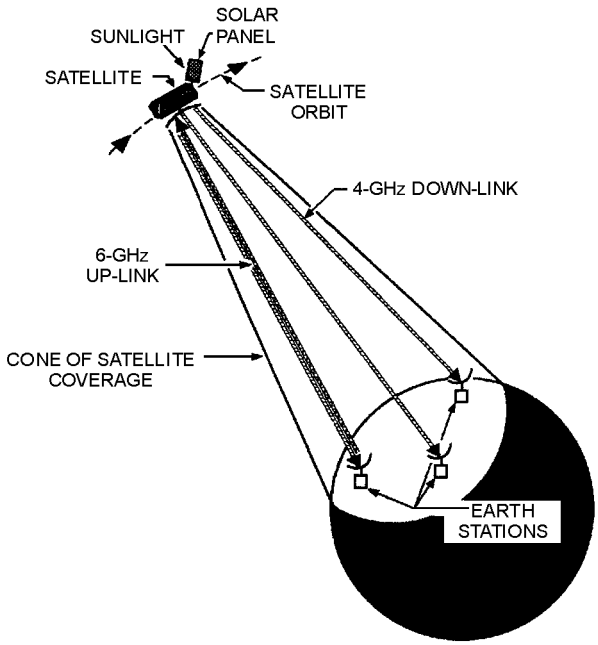


Figure 4-1.—Satellite/earth station communications system.

A tremendous amount of knowledge and information has been gained about the design of antennas and radio-wave propagation. Still, many old-time technicians will tell you that when it comes to designing the length of an antenna, the best procedure is to perform all calculations and try out the antenna. If it doesn't work right, use a cut-and-try method until it does. Fortunately, enough information has been collected over the last few decades that it is now possible to predict the behavior of antennas. This chapter will discuss and explain the basic design and operation of antennas.

PRINCIPLES OF ANTENNA RADIATION

After an rf signal has been generated in a transmitter, some means must be used to radiate this signal through space to a receiver. The device that does this job is the antenna. The transmitter signal energy is sent into space by a TRANSMITTING ANTENNA; the rf signal is then picked up from space by a RECEIVING ANTENNA.

The rf energy is transmitted into space in the form of an electromagnetic field. As the traveling electromagnetic field arrives at the receiving antenna, a voltage is induced into the antenna (a conductor). The rf voltages induced into the receiving antenna are then passed into the receiver and converted back into the transmitted rf information.

The design of the antenna system is very important in a transmitting station. The antenna must be able to radiate efficiently so the power supplied by the transmitter is not wasted. An efficient transmitting antenna must have exact dimensions. The dimensions are determined by the transmitting frequencies. The dimensions of the receiving antenna are not critical for relatively low radio frequencies. However, as the frequency of the signal being received increases, the design and installation of the receiving antenna become more critical. An example of this is a television receiving antenna. If you raise it a few more inches from the ground or give a slight turn in direction, you can change a snowy blur into a clear picture.

The conventional antenna is a conductor, or system of conductors, that radiates or intercepts electromagnetic wave energy. An ideal antenna has a definite length and a uniform diameter, and is completely isolated in space. However, this ideal antenna is not realistic. Many factors make the design of an antenna for a communications system a more complex problem than you would expect. These factors include the height of the radiator above the earth, the conductivity of the earth below it, and the shape and dimensions of the antenna. All of these factors affect the radiated-field pattern of the antenna in space. Another problem in antenna design is that the radiation pattern of the antenna must be directed between certain angles in a horizontal or vertical plane, or both.

Most practical transmitting antennas are divided into two basic classifications, HERTZ (half-wave) ANTENNAS and MARCONI (quarter-wave) ANTENNAS. Hertz antennas are generally installed some distance above the ground and are positioned to radiate either vertically or horizontally. Marconi antennas operate with one end grounded and are mounted perpendicular to the Earth or to a surface acting as a ground. Hertz antennas are generally used for frequencies above 2 megahertz. Marconi antennas are used for frequencies below 2 megahertz and may be used at higher frequencies in certain applications.

A complete antenna system consists of three parts: (1) The COUPLING DEVICE, (2) the FEEDER, and (3) the ANTENNA, as shown in figure 4-2. The coupling device (coupling coil) connects the transmitter to the feeder. The feeder is a transmission line that carries energy to the antenna. The antenna radiates this energy into space.

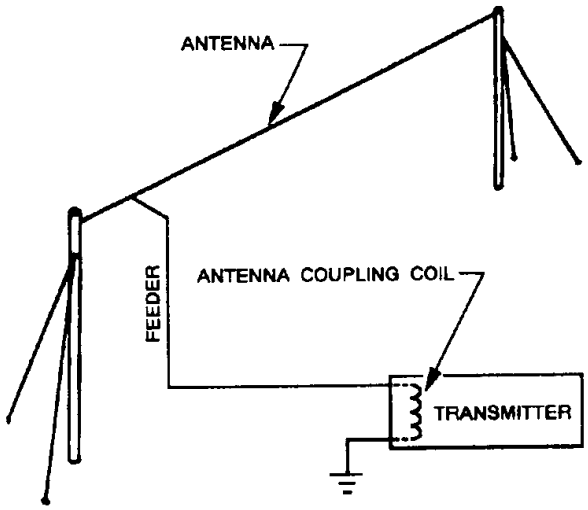


Figure 4-2.—Typical antenna system.

The factors that determine the type, size, and shape of the antenna are (1) the frequency of operation of the transmitter, (2) the amount of power to be radiated, and (3) the general direction of the receiving set. Typical antennas are shown in figure 4-3.

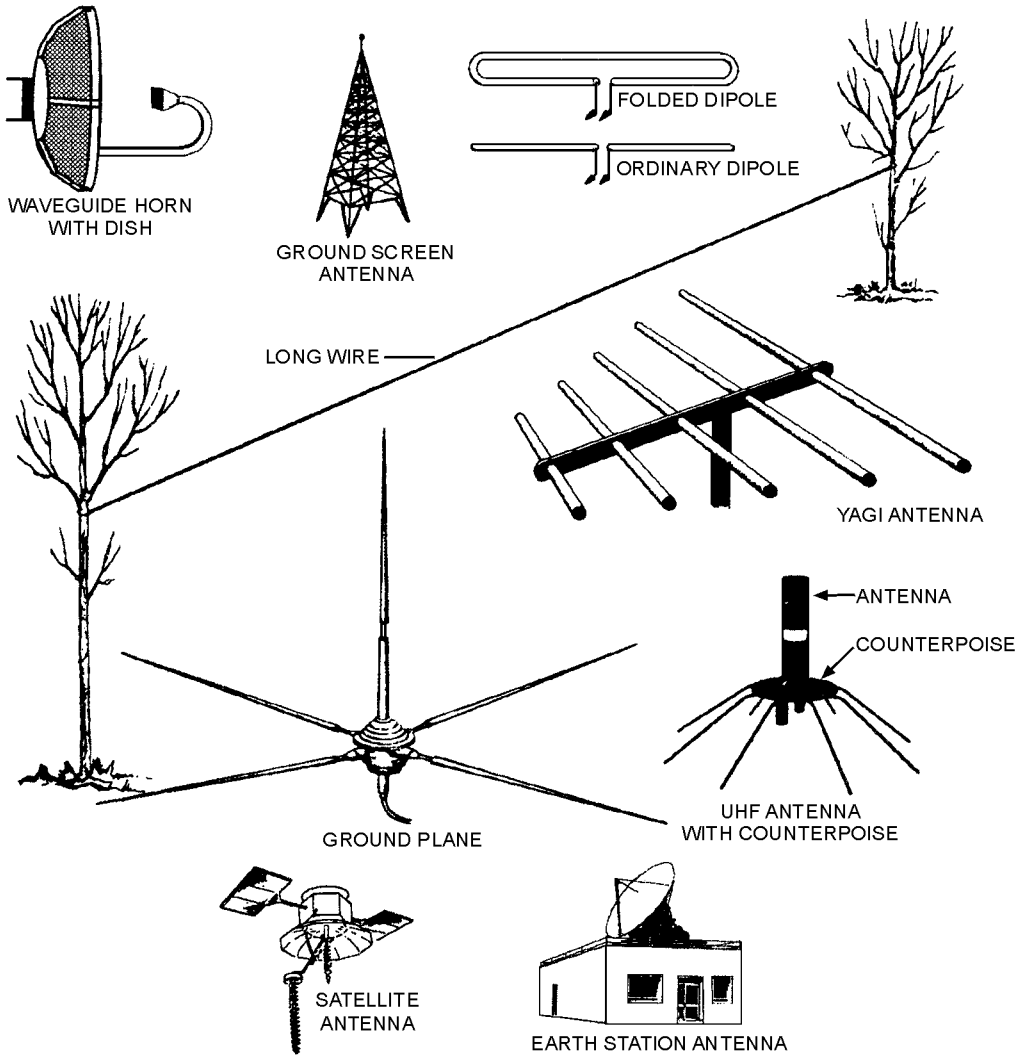


Figure 4-3.—Typical antennas.

CURRENT AND VOLTAGE DISTRIBUTION ON AN ANTENNA

A current flowing in a wire whose length is properly related to the rf produces an electro magnetic field. This field is radiated from the wire and is set free in space. We will discuss how these waves are set free later in this chapter. Remember, the principles of radiation of electromagnetic energy are based on two laws:

1. A MOVING ELECTRIC FIELD CREATES A MAGNETIC (H) FIELD.

2. A MOVING MAGNETIC FIELD CREATES AN ELECTRIC (E) FIELD.

In space, these two fields will be in phase and perpendicular to each other at any given time. Although a conductor is usually considered present when a moving electric or magnetic field is mentioned, the laws that govern these fields say nothing about a conductor. Therefore, these laws hold true whether a conductor is present or not.

Figure 4-4 shows the current and voltage distribution on a half-wave (Hertz) antenna. In view A, a piece of wire is cut in half and attached to the terminals of a high-frequency ac generator. The frequency of the generator is set so that each half of the wire is 1/4 wavelength of the output. The result is a common type of antenna known as a DIPOLE.

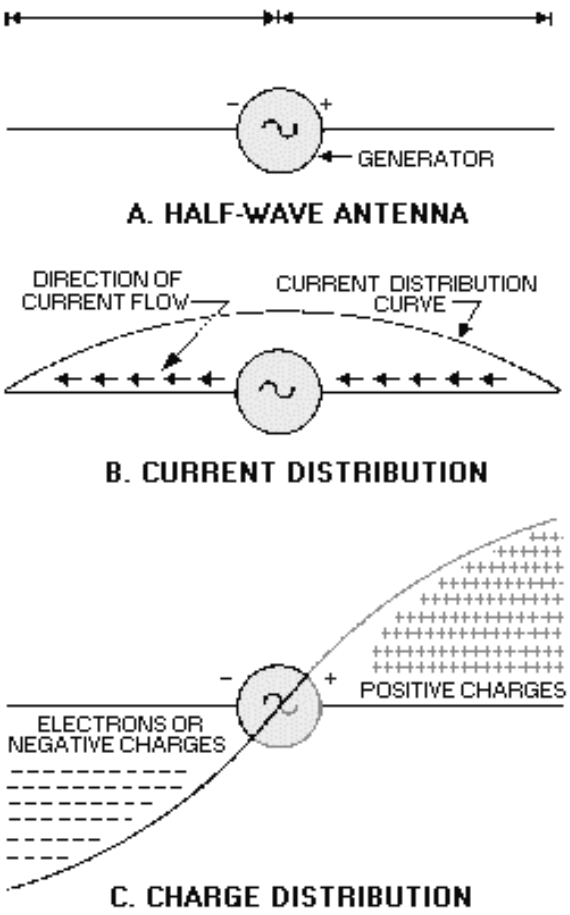


Figure 4-4.—Current and voltage distribution on an antenna.

At a given time the right side of the generator is positive and the left side negative. Remember that like charges repel. Because of this, electrons will flow away from the negative terminal as far as possible, but will be attracted to the positive terminal. View B shows the direction and distribution of electron flow. The distribution curve shows that most current flows in the center and none flows at the ends. The current distribution over the antenna will always be the same no matter how much or how little current is flowing. However, current at any given point on the antenna will vary directly with the amount of voltage developed by the generator.

One-quarter cycle after electrons have begun to flow, the generator will develop its maximum voltage and the current will decrease to 0. At that time the condition shown in view C will exist. No current will be flowing, but a maximum number of electrons will be at the left end of the line and a minimum number at the right end. The charge distribution view C along the wire will vary as the voltage of the generator varies. Therefore, you may draw the following conclusions:

1. A current flows in the antenna with an amplitude that varies with the generator voltage.

2. A sinusoidal distribution of charge exists on the antenna. Every 1/2 cycle, the charges reverse polarity.

3. The sinusoidal variation in charge magnitude lags the sinusoidal variation in current by 1/4 cycle.

Q1. What are the two basic classifications of antennas?

Q2. What are the three parts of a complete antenna system?

Q3. What three factors determine the type, size, and shape of an antenna?

RADIATION OF ELECTROMAGNETIC ENERGY

The electromagnetic radiation from an antenna is made up of two components, the E field and the H field. We discussed these fields in chapters 1 and 2. The two fields occur 90 degrees out of phase with each other. These fields add and produce a single electromagnetic field. The total energy in the radiated wave remains constant in space except for some absorption of energy by the Earth. However, as the wave advances, the energy spreads out over a greater area and, at any given point, decreases as the distance increases.

Various factors in the antenna circuit affect the radiation of these waves. In figure 4-5, for example, if an alternating current is applied at the A end of the length of wire from A to B, the wave will travel along the wire until it reaches the B end. Since the B end is free, an open circuit exists and the wave cannot travel farther. This is a point of high impedance. The wave bounces back (reflects) from this point of high impedance and travels toward the starting point, where it is again reflected. The energy of the wave would be gradually dissipated by the resistance of the wire of this back-and-forth motion (oscillation); however, each time it reaches the starting point, the wave is reinforced by an amount sufficient to replace the energy lost. This results in continuous oscillations of energy along the wire and a high voltage at the A end of the wire. These oscillations are applied to the antenna at a rate equal to the frequency of the rf voltage.

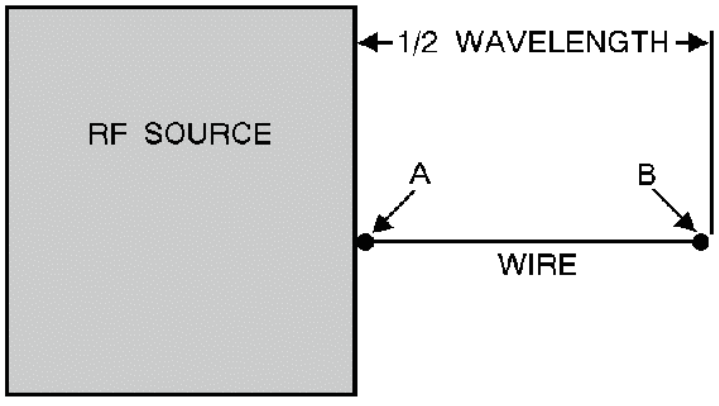


Figure 4-5.—Antenna and rf source.

These impulses must be properly timed to sustain oscillations in the antenna. The rate at which the waves travel along the wire is constant at approximately 300,000,000 meters per second. The length of

the antenna must be such that a wave will travel from one end to the other and back again during the period of 1 cycle of the rf voltage. Remember, the distance a wave travels during the period of 1 cycle is known as the wavelength and is found by dividing the rate of travel by the frequency.

Look at the current and voltage (charge) distribution on the antenna in figure 4-6. A maximum movement of electrons is in the center of the antenna at all times; therefore, the center of the antenna is at a low impedance. This condition is called a STANDING WAVE of current. The points of high current and high voltage are known as current and voltage LOOPS. The points of minimum current and minimum voltage are known as current and voltage NODES. View A shows a current loop and current nodes. View B shows voltage loops and a voltage node. View C shows the resultant voltage and current loops and nodes. The presence of standing waves describes the condition of resonance in an antenna. At resonance the waves travel back and forth in the antenna reinforcing each other and the electromagnetic waves are transmitted into space at maximum radiation. When the antenna is not at resonance, the waves tend to cancel each other and lose energy in the form of heat.

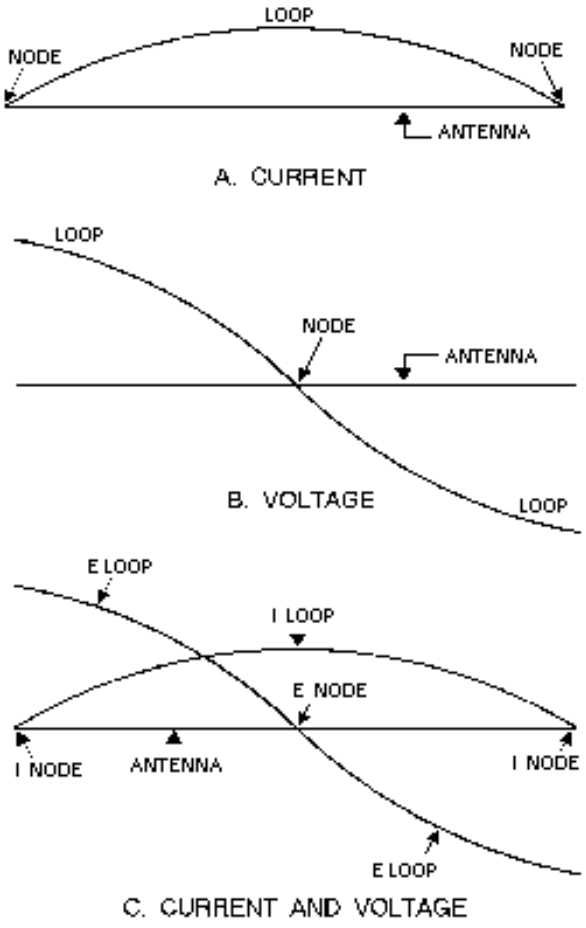


Figure 4-6.—Standing waves of voltage and current on an antenna.

Q4. If a wave travels exactly the length of an antenna from one end to the other and back during the period of 1 cycle, what is the length of the antenna?

Q5. What is the term used to identify the points of high current and high voltage on an antenna?

Q6. What is the term used to identify the points of minimum current and minimum voltage on an antenna?

ANTENNA CHARACTERISTICS

You can define an antenna as a conductor or group of conductors used either for radiating electromagnetic energy into space or for collecting it from space. Electrical energy from the transmitter is converted into electromagnetic energy by the antenna and radiated into space. On the receiving end, electromagnetic energy is converted into electrical energy by the antenna and is fed into the receiver.

Fortunately, separate antennas seldom are required for both transmitting and receiving rf energy. Any antenna can transfer energy from space to its input receiver with the same efficiency that it transfers energy from the transmitter into space. Of course, this is assuming that the same frequency is used in both cases. This property of interchangeability of the same antenna for transmitting and receiving is known as antenna RECIPROCITY. Antenna reciprocity is possible because antenna characteristics are essentially the same for sending and receiving electromagnetic energy.

RECIPROCITY OF ANTENNAS

In general, the various properties of an antenna apply equally, regardless of whether you use the antenna for transmitting or receiving. The more efficient a certain antenna is for transmitting, the more efficient it will be for receiving on the same frequency. Likewise, the directive properties of a given antenna also will be the same whether it is used for transmitting or receiving.

Assume, for example, that a certain antenna used with a transmitter radiates a maximum amount of energy at right angles to the axis of the antenna, as shown in figure 4-7, view A. Note the minimum amount of radiation along the axis of the antenna. Now, if this same antenna were used as a receiving antenna, as shown in view B, it would receive best in the same directions in which it produced maximum radiation; that is, at right angles to the axis of the antenna.

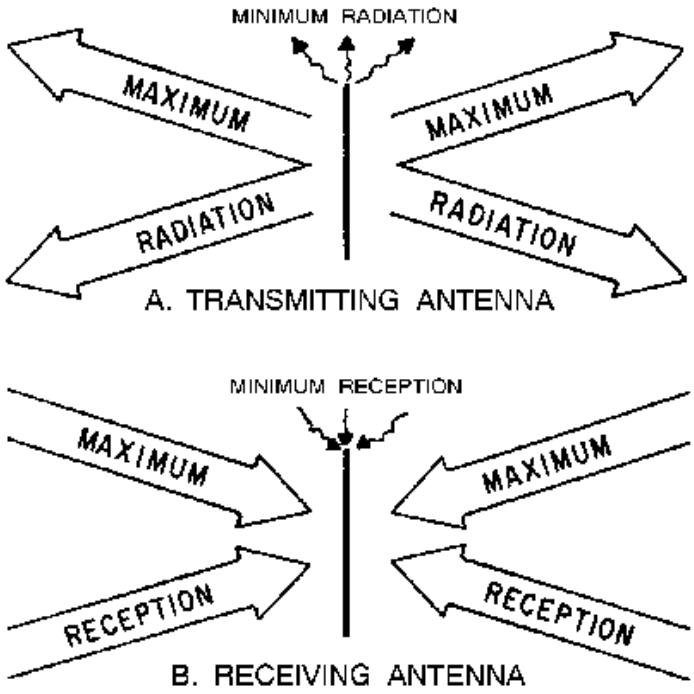


Figure 4-7.—Reciprocity of antennas.

ANTENNA GAIN

Another characteristic of a given antenna that remains the same whether the antenna is used for transmitting or receiving is GAIN. Some antennas are highly directional that is, more energy is propagated in certain directions than in others. The ratio between the amount of energy propagated in these directions compared to the energy that would be propagated if the antenna were not directional is known as its gain. When a transmitting antenna with a certain gain is used as a receiving antenna, it will also have the same gain for receiving.

POLARIZATION

Let's review polarization briefly. In chapter 2 you learned that the radiation field is composed of electric and magnetic lines of force. These lines of force are always at right angles to each other. Their intensities rise and fall together, reaching their maximums 90 degrees apart. The electric field determines the direction of polarization of the wave. In a vertically polarized wave, the electric lines of force lie in a vertical direction. In a horizontally polarized wave, the electric lines of force lie in a horizontal direction. Circular polarization has the electric lines of force rotating through 360 degrees with every cycle of rf energy.

The electric field was chosen as the reference field because the intensity of the wave is usually measured in terms of the electric field intensity (volts, millivolts, or microvolts per meter). When a single-wire antenna is used to extract energy from a passing radio wave, maximum pickup will result when the antenna is oriented in the same direction as the electric field. Thus a vertical antenna is used for the efficient reception of vertically polarized waves, and a horizontal antenna is used for the reception of horizontally polarized waves. In some cases the orientation of the electric field does not remain constant.

Instead, the field rotates as the wave travels through space. Under these conditions both horizontal and vertical components of the field exist and the wave is said to have an elliptical polarization.

Q7. The various properties of a transmitting antenna can apply equally to the same antenna when it is used as a receiving antenna. What term is used for this property?

Q8. The direction of what field is used to designate the polarization of a wave?

Q9. If a wave's electric lines of force rotate through 360 degrees with every cycle of rf energy, what is the polarization of this wave?

Polarization Requirements for Various Frequencies

Ground-wave transmission is widely used at medium and low frequencies. Horizontal polarization cannot be used at these frequencies because the electric lines of force are parallel to and touch the earth. Since the earth acts as a fairly good conductor at low frequencies, it would short out the horizontal electric lines of force and prevent the radio wave from traveling very far. Vertical electric lines of force, on the other hand, are bothered very little by the earth. Therefore vertical polarization is used for ground-wave transmission, allowing the radio wave to travel a considerable distance along the ground surface with minimum attenuation.

Sky-wave transmission is used at high frequencies. Either horizontal or vertical polarization can be used with sky-wave transmission because the sky wave arrives at the receiving antenna elliptically polarized. This is the result of the wave traveling obliquely through the Earth's magnetic field and striking the ionosphere. The radio wave is given a twisting motion as it strikes the ionosphere. Its orientation continues to change because of the unstable nature of the ionosphere. The relative amplitudes and phase differences between the horizontal and vertical components of the received wave also change. Therefore, the transmitting and receiving antennas can be mounted either horizontally or vertically.

Although either horizontally or vertically polarized antennas can be used for high frequencies, horizontally polarized antennas have certain advantages and are therefore preferred. One advantage is that vertically polarized interference signals, such as those produced by automobile ignition systems and electrical appliances, are minimized by horizontal polarization. Also, less absorption of radiated energy by buildings or wiring occurs when these antennas are used. Another advantage is that support structures for these antennas are of more convenient size than those for vertically polarized antennas.

For frequencies in the vhf or uhf range, either horizontal or vertical polarization is satisfactory. These radio waves travel directly from the transmitting antenna to the receiving antenna without entering the ionosphere. The original polarization produced at the transmitting antenna is maintained throughout the entire travel of the wave to the receiver. Therefore, if a horizontally polarized antenna is used for transmitting, a horizontally polarized antenna must be used for receiving. The requirements would be the same for a vertical transmitting and receiving antenna system.

For satellite communications, parallel frequencies can be used without interference by using polarized radiation. The system setup is shown in figure 4-8. One pair of satellite antennas is vertically polarized and another pair is horizontally polarized. Either vertically or horizontally polarized transmissions are received by the respective antenna and retransmitted in the same polarization. For example, transmissions may be made in the 3.7 to 3.74 GHz range on the vertical polarization path and in the 3.72 to 3.76 GHz range on the horizontal polarization path without adjacent frequency (co-channel) interference.

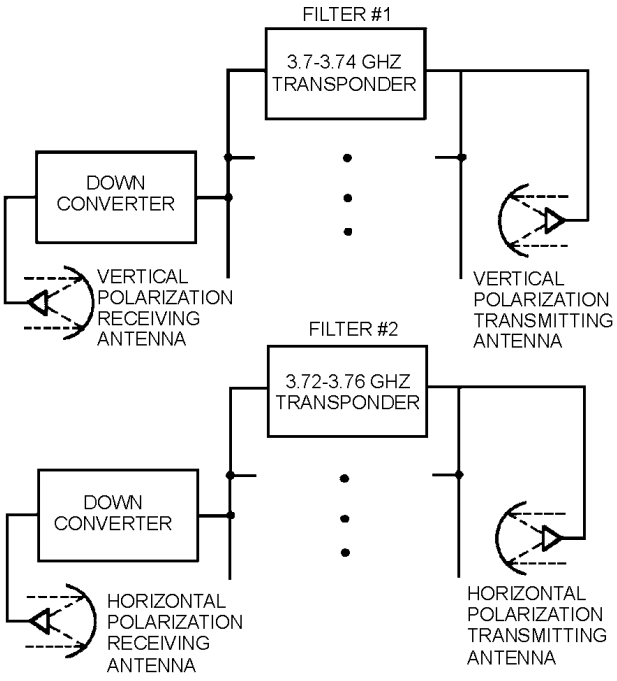


Figure 4-8.—Satellite transmissions using polarized radiation.

Advantages of Vertical Polarization

Simple vertical antennas can be used to provide OMNIDIRECTIONAL (all directions) communication. This is an advantage when communications must take place from a moving vehicle.

In some overland communications, such as in vehicular installations, antenna heights are limited to 3 meters (10 feet) or less. In such instances vertical polarization results in a stronger receiver signal than does horizontal polarization at frequencies up to about 50 megahertz. From approximately 50 to 100 megahertz, vertical polarization results in a slightly stronger signal than does horizontal polarization with antennas at the same height. Above 100 megahertz, the difference in signal strength is negligible.

For transmission over bodies of water, vertical polarization is much better than horizontal polarization for antennas at the lower heights. As the frequency increases, the minimum antenna height decreases. At 30 megahertz, vertical polarization is better for antenna heights below about 91 meters (300 feet); at 85 megahertz, antenna heights below 15 meters (50 feet); and still lower heights at the high frequencies. Therefore, at ordinary antenna mast heights of 12 meters (40 feet), vertical polarization is advantageous for frequencies less than about 100 megahertz.

Radiation is somewhat less affected by reflections from aircraft flying over the transmission path when vertical polarization is used instead of horizontal polarization. With horizontal polarization, such reflections cause variations in received signal strength. This factor is important in locations where aircraft traffic is heavy.

When vertical polarization is used, less interference is produced or picked up because of strong vhf and uhf broadcast transmissions (television and fm). This is because vhf and uhf transmissions use horizontal polarization. This factor is important when an antenna must be located in an urban area having several television and fm broadcast stations.

Advantages of Horizontal Polarization

A simple horizontal antenna is bi-directional. This characteristic is useful when you desire to minimize interference from certain directions. Horizontal antennas are less likely to pick up man-made interference, which ordinarily is vertically polarized.

When antennas are located near dense forests or among buildings, horizontally polarized waves suffer lower losses than vertically polarized waves, especially above 100 megahertz. Small changes in antenna locations do not cause large variations in the field intensity of horizontally polarized waves. When vertical polarization is used, a change of only a few meters in the antenna location may have a considerable effect on the received signal strength. This is the result of interference patterns that produce standing waves in space when spurious reflections from trees or buildings occur.

When simple antennas are used, the transmission line, which is usually vertical, is less affected by a horizontally mounted antenna. When the antenna is mounted at right angles to the transmission line and horizontal polarization is used, the line is kept out of the direct field of the antenna. As a result, the radiation pattern and electrical characteristics of the antenna are practically unaffected by the presence of the vertical transmission line.

Q10. What type of polarization should be used at medium and low frequencies?

Q11. What is an advantage of using horizontal polarization at high frequencies?

Q12. What type of polarization should be used if an antenna is mounted on a moving vehicle at frequencies below 50 megahertz?

RADIATION RESISTANCE

Radiated energy is the useful part of the transmitter's signal. However, it represents as much of a loss to the antenna as the energy lost in heating the antenna wire. In either case, the dissipated power is equal to I^2R. In the case of heat losses, the R is real resistance. In the case of radiation, R is an assumed resistance; if this resistance were actually present, it would dissipate the same amount of power that the antenna takes to radiate the energy. This assumed resistance is referred to as the RADIATION RESISTANCE.

Radiation resistance varies at different points on the antenna. This resistance is always measured at a current loop. For the antenna in free space, that is, entirely removed from any objects that might affect its operation, the radiation resistance is 73 ohms. A practical antenna located over a ground plane may have any value of radiation resistance from 0 to approximately 100 ohms. The exact value of radiation resistance depends on the height of the antenna above the ground. For most half-wave wire antennas, the radiation resistance is about 65 ohms. It will usually vary between 55 and 600 ohms for antennas constructed of rod or tubing. The actual value of radiation resistance, so long as it is 50 ohms or more, has little effect on the radiation efficiency of the antenna. This is because the ohmic resistance is about 1 ohm for conductors of large diameter. The ohmic resistance does not become important until the radiation resistance drops to a value less than 10 ohms. This may be the case when several antennas are coupled together.

RADIATION TYPES AND PATTERNS

The energy radiated from an antenna forms a field having a definite RADIATION PATTERN. A radiation pattern is a plot of the radiated energy from an antenna. This energy is measured at various angles at a constant distance from the antenna. The shape of this pattern depends on the type of antenna

used. In this section, we will introduce the basic types of radiation (isotropic and anisotropic) and their radiation patterns.

Isotropic Radiation

Some antenna sources radiate energy equally in all directions. Radiation of this type is known as ISOTROPIC RADIATION. We all know the Sun radiates energy in all directions. The energy radiated from the Sun measured at any fixed distance and from any angle will be approximately the same. Assume that a measuring device is moved around the Sun and stopped at the points indicated in figure 4-9 to make a measurement of the amount of radiation. At any point around the circle, the distance from the measuring device to the Sun is the same. The measured radiation will also be the same. The Sun is therefore considered an isotropic radiator.

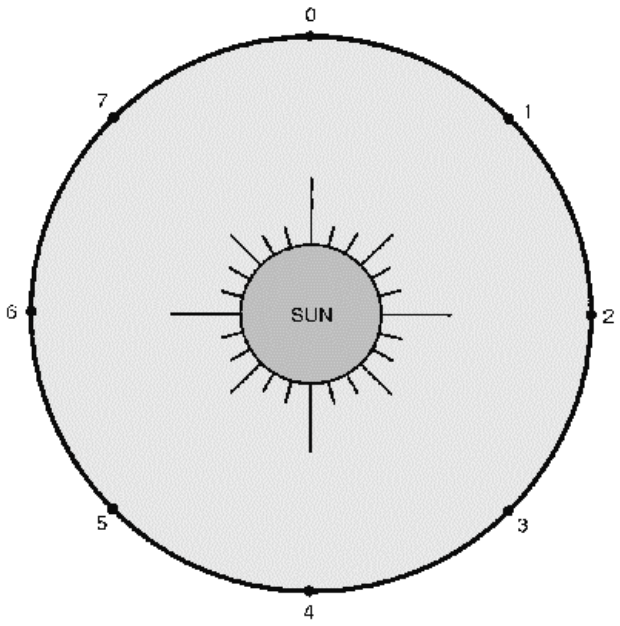


Figure 4-9.—Isotropic radiator.

To plot this pattern, we will assume that the radiation is measured on a scale of 0 to 10 units and that the measured amount of radiation is 7 units at all points. We will then plot our measurements on two different types of graphs, rectangular- and polar-coordinate graphs. The RECTANGULAR-COORDINATE GRAPH of the measured radiation, shown in view A of figure 4-10, is a straight line plotted against positions along the circle. View B shows the POLAR-COORDINATE GRAPH for the same isotropic source.

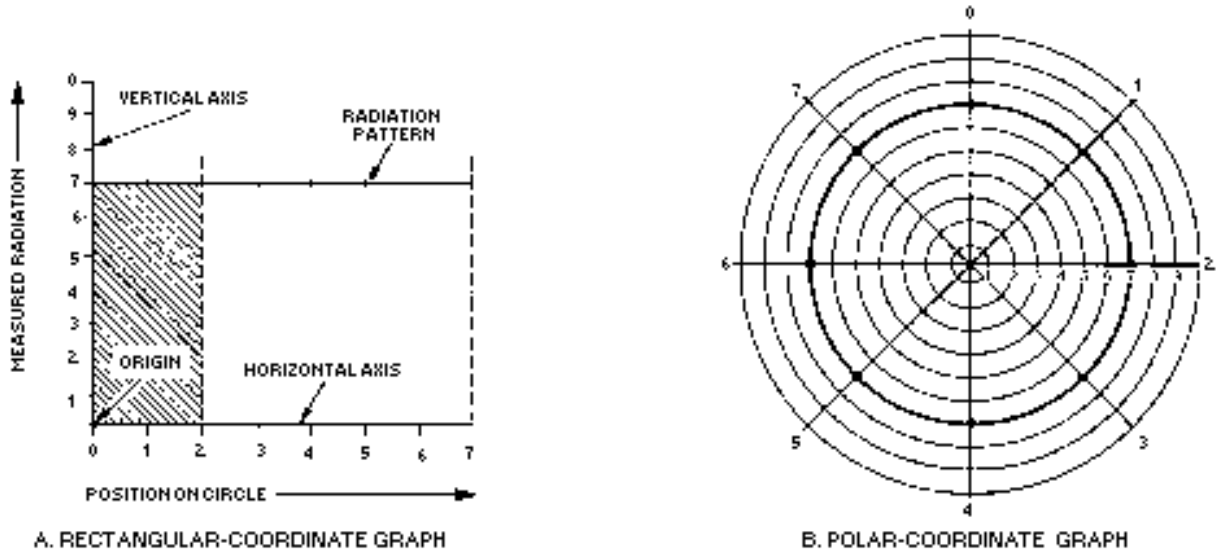


Figure 4-10.—Comparison of rectangular- and polar-coordinate graphs for an isotropic source.

In the rectangular-coordinate graph, points are located by projection from a pair of stationary, perpendicular axes. In the polar-coordinate graph, points are located by projection along a rotating axis (radius) to an intersection with one of several concentric, equally-spaced circles. The horizontal axis on the rectangular-coordinate graph corresponds to the circles on the polar-coordinate graph. The vertical axis on the rectangular-coordinate graph corresponds to the rotating axis (radius) on the polar-coordinate graph.

Rectangular-Coordinate Pattern

Look at view A of figure 4-10. The numbered positions around the circle are laid out on the HORIZONTAL AXIS of the graph from 0 to 7 units. The measured radiation is laid out on the VERTICAL AXIS of the graph from 0 to 10 units. The units on both axes are chosen so the pattern occupies a convenient part of the graph.

The horizontal and vertical axes are at a right angle to each other. The point where the axes cross each other is known as the ORIGIN. In this case, the origin is 0 on both axes. Now, assume that a radiation value of 7 units view B is measured at position 2. From position 2 on the horizontal axis, a dotted line is projected upwards that runs parallel to the vertical axis. From position 7 on the vertical axis, a line is projected to the right that runs parallel to the horizontal axis. The point where the two lines cross (INTERCEPT) represents a value of 7 radiation units at position 2. This is the only point on the graph that can represent this value.

As you can see from the figure, the lines used to plot the point form a rectangle. For this reason, this type of plot is called a *rectangular-coordinate graph*. A new rectangle is formed for each different point plotted. In this example, the points plotted lie in a straight line extending from 7 units on the vertical scale to the projection of position 7 on the horizontal scale. This is the characteristic pattern in rectangular coordinates of an isotropic source of radiation.

Polar-Coordinate Pattern

The polar-coordinate graph has proved to be of great use in studying radiation patterns. Compare views A and B of figure 4-10. Note the great difference in the shape of the radiation pattern when it is

transferred from the rectangular-coordinate graph in view A to the polar-coordinate graph in view B. The scale of radiation values used in both graphs is identical, and the measurements taken are both the same. However, the shape of the pattern is drastically different.

Look at view B of figure 4-10 and assume that the center of the concentric circles is the Sun. Assume that a radius is drawn from the Sun (center of the circle) to position 0 of the circle. When you move to position 1, the radius moves to position 1; when you move to position 2, the radius also moves to position 2, and so on.

The positions where a measurement was taken are marked as 0 through 7 on the graph. Note how the position of the radius indicates the actual direction from the source at which the measurement was taken. This is a distinct advantage over the rectangular-coordinate graph in which the position is indicated along a straight-line axis and has no physical relation to the actual direction of measurement. Now that we have a way to indicate the *direction* of measurement, we must devise a way to indicate the *magnitude* of the radiation.

Notice that the rotating axis is always drawn from the center of the graph to some position on the edge of the graph. As the axis moves toward the edge of the graph, it passes through a set of equally-spaced, concentric circles. In this example view B, they are numbered successively from 1 to 10 from the center out. These circles are used to indicate the magnitude of the radiation.

The advantages of the polar-coordinate graph are immediately evident. The source, which is at the center of the observation circles, is also at the center of the graph. By looking at a polar-coordinate plot of a radiation pattern, you can immediately see the direction and strength of radiation put out by the source. Therefore, the polar-coordinate graph is more useful than the rectangular-coordinate graph in plotting radiation patterns.

Anisotropic Radiation

Most radiators emit (radiate) stronger radiation in one direction than in another. A radiator such as this is referred to as ANISOTROPIC. An example of an anisotropic radiator is an ordinary flashlight. The beam of the flashlight lights only a portion of the space surrounding it. If a circle is drawn with the flashlight as the center, as shown in view B of figure 4-11, the radiated light can be measured at different positions around the circle. Again, as with the isotropic radiator, all positions are the same distance from the center, but at different angles. However, in this illustration the radiated light is measured at 16 different positions on the circle.

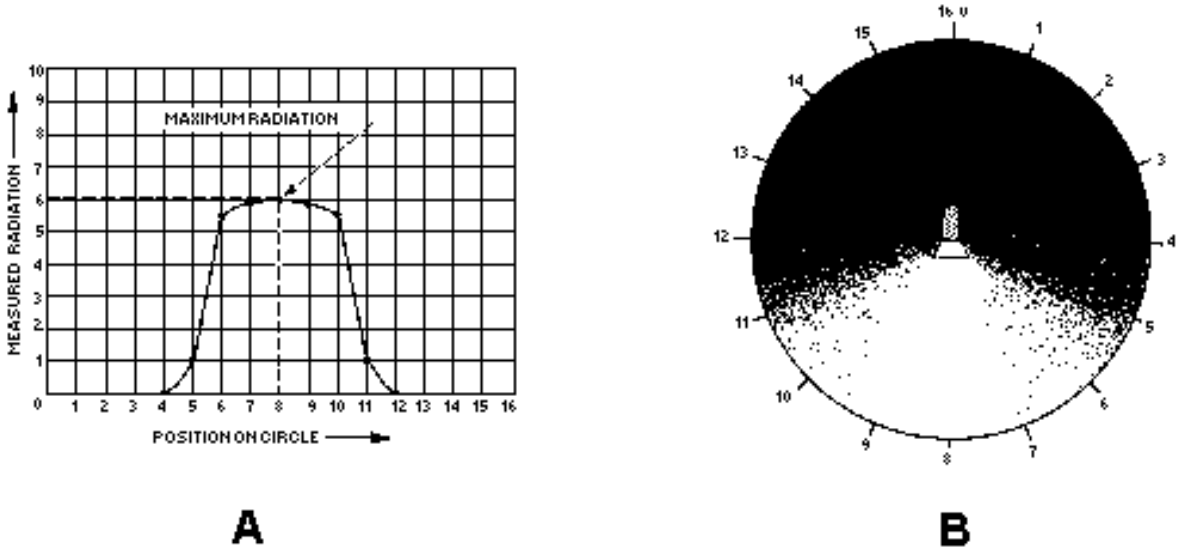


Figure 4-11.—Anisotropic radiator.

Directly behind the flashlight (position 0) the radiation measured is minimum. Accordingly, a 0 value is assigned to this position in the rectangular-coordinate graph (fig. 4-11, view A). This radiation remains at minimum until position 4 is reached. Between positions 4 and 6, the measuring device enters the flashlight beam. You can see this transition from darkness to brightness easily in view B. Radiation is fairly constant between positions 6 and 10. Maximum brightness occurs at position 8, which is directly in the path of the flashlight beam. From positions 10 to 12, the measuring device leaves the flashlight beam and the radiation measurement falls off sharply. At position 13 the radiation is again at 0 and stays at this value back to position 0.

Radiation from a light source and radiation from an antenna are both forms of electromagnetic waves. Therefore, the measurement of radiation of an antenna follows the same basic procedure as that just described for the Sun and the flashlight. Before proceeding further with the study of antenna patterns, you should be sure you understand the methods used to graph the measured radiation (magnitude of the radiation). Study the rectangular- and polar-coordinate systems of plotting presented in the following section.

Q13. What is the radiation resistance of a half-wave antenna in free space?

Q14. A radiating source that radiates energy stronger in one direction than another is known as what type of radiator?

Q15. A radiating source that radiates energy equally in all directions is known as what type of radiator?

Q16. A flashlight is an example of what type of radiator?

In figure 4-11, view A, the radiation pattern of the flashlight is graphed in rectangular coordinates. The illustration of the flashlight beam in view B clearly indicates the shape of the flashlight beam. This is not evident in the radiation pattern plotted on the rectangular-coordinate graph. Now look at figure 4-12. The radiation pattern shown in this figure looks very much like the actual flashlight beam. The pattern in figure 4-12 is plotted using the same values as those of figure 4-11, view A, but is drawn using polar coordinates.

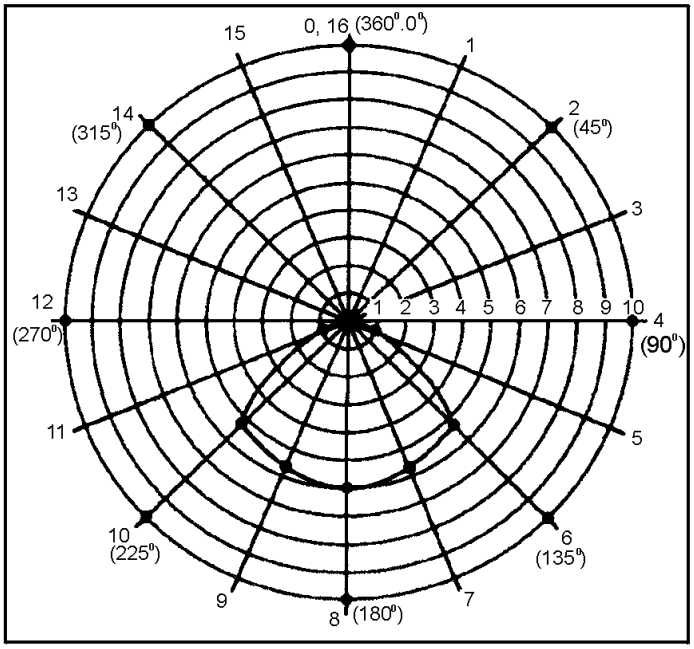


Figure 4-12.—Polar-coordinate graph for anisotropic radiator.

The positions marked off on the two polar-coordinate graphs in figures 4-10 and 4-12 were selected and numbered arbitrarily. However, a standard method allows the positions around a source to be marked off so that one radiation pattern can easily be compared with another. This method is based on the fact that a circle has a radius of 360 degrees. The radius extending vertically from the center (position 0 in figure 4-10) is designated 0 degrees. At position 4 the radius is at a right angle to the 0-degree radius. Accordingly, the radius at position 4 is marked 90 degrees, position 8 is 180 degrees, position 12 is 270 degrees, and position 16 is 360 degrees. The various radii drawn on the graph are all marked according to the angle each radius makes with the reference radius at 0 degrees.

The radiation pattern in figure 4-12 is obtained by using the same procedure that was used for (figure 4-10, view B). The radiation measured at positions 1, 2, 3, and 4 is 0. Position 5 measures approximately 1 unit. This is marked on the graph and the rotating radius moves to position 6. At this position a reading of 5.5 units is taken. As before, this point is marked on the graph. The procedure is repeated around the circle and a reading is obtained from positions 6 through 11. At position 12 no radiation is indicated, and this continues on to position 16.

The polar-coordinate graph now shows a definite area enclosed by the radiation pattern. This pattern indicates the general direction of radiation from the source. The enclosed area is called a LOBE. Outside of this area, minimum radiation is emitted in any direction. For example, at position 2 the radiation is 0. Such a point is called a NULL. In real situations, some radiation is usually transmitted in all directions. Therefore, a null is used to indicate directions of minimum radiation. The pattern of figure 4-12 shows one lobe and one continuous null.

ANTENNA LOADING

You will sometimes want to use one antenna system for transmitting and receiving on several different frequencies. Since the antenna must always be in resonance with the applied frequency, you may need to either physically or electrically lengthen or shorten the antenna.

Except for trailing-wire antennas used in aircraft installations (which may be lengthened or shortened), physically lengthening the antenna is not very practical. But you can achieve the same result by changing the electrical length of the antenna. To change the electrical length, you can insert either an inductor or a capacitor in series with the antenna. This is shown in figure 4-13, views A and B. Changing the electrical length by this method is known as LUMPED-IMPEDANCE TUNING, or LOADING. The electrical length of any antenna wire can be increased or decreased by loading. If the antenna is too short for the wavelength being used, it is resonant at a higher frequency than that at which it is being excited. Therefore, it offers a capacitive reactance at the excitation frequency. This capacitive reactance can be compensated for by introducing a lumped-inductive reactance, as shown in view A. Similarly, if the antenna is too long for the transmitting frequency, it offers an inductive reactance. Inductive reactance can be compensated for by introducing a lumped-capacitive reactance, as shown in view B. An antenna without loading is represented in view C.

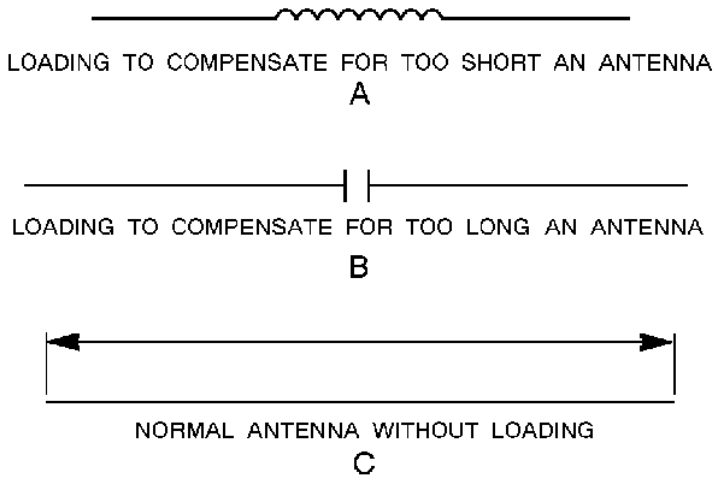


Figure 4-13.—Electrically equal antenna.

BASIC ANTENNAS

Before you look at the various types of antennas, consider the relationship between the wavelength at which the antenna is being operated and the actual length of the antenna. An antenna does not necessarily radiate or receive more energy when it is made longer. Specific dimensions must be used for efficient antenna operation.

Nearly all antennas have been developed from two basic types, the Hertz and the Marconi. The basic Hertz antenna is 1/2 wavelength long at the operating frequency and is insulated from ground. It is often called a DIPOLE or a DOUBLET. The basic Marconi antenna is 1/4 wavelength long and is either grounded at one end or connected to a network of wires called a COUNTERPOISE. The ground or counterpoise provides the equivalent of an additional 1/4 wavelength, which is required for the antenna to resonate.

HALF-WAVE ANTENNAS

A half-wave antenna (referred to as a dipole, Hertz, or doublet) consists of two lengths of wire rod, or tubing, each 1/4 wavelength long at a certain frequency. It is the basic unit from which many complex antennas are constructed. The half-wave antenna operates independently of ground; therefore, it may be installed far above the surface of the Earth or other absorbing bodies. For a dipole, the current is

maximum at the center and minimum at the ends. Voltage is minimum at the center and maximum at the ends, as was shown in figure 4-6.

Radiation Patterns

In the following discussion, the term DIPOLE is used to mean the basic half-wave antenna. The term DOUBLET is used to indicate an antenna that is very short compared with the wavelength of the operating frequency. Physically, it has the same shape as the dipole.

RADIATION PATTERN OF A DOUBLET.—The doublet is the simplest form of a practical antenna. Its radiation pattern can be plotted like the radiation pattern of the flashlight (fig. 4-12). Figure 4-14 shows the development of vertical and horizontal patterns for a doublet. This in NOT a picture of the radiation, but three-dimensional views of the pattern itself. In three views the pattern resembles a doughnut. From the dimensions in these views, two types of polar-coordinate patterns can be drawn, horizontal and vertical. The HORIZONTAL PATTERN view A is derived from the solid pattern view C by slicing it horizontally. This produces view B, which is converted to the polar coordinates seen in view A. The horizontal pattern illustrates that the radiation is constant in any direction along the horizontal plane.

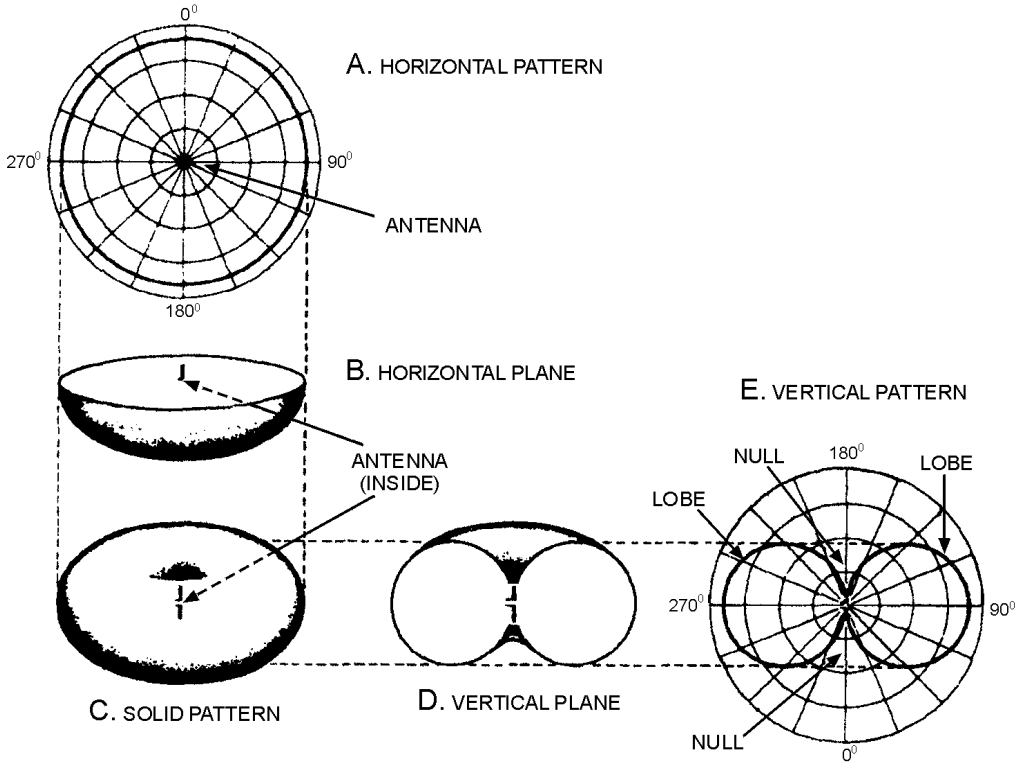


Figure 4-14.—Development of vertical and horizontal patterns.

A VERTICAL PATTERN view E is obtained from the drawing of the vertical plane view D of the radiation pattern view C. The radiation pattern view C is sliced in half along a vertical plane through the antenna. This produces the vertical plane pattern in view D. Note how the vertical plane in view D of the radiation pattern differs from the horizontal plane in view B. The vertical pattern view E exhibits two lobes and two nulls. The difference between the two patterns is caused by two facts: (1) no radiation is

emitted from the ends of the doublet; and (2) maximum radiation comes from the doublet in a direction perpendicular to the antenna axis. This type of radiation pattern is both NONDIRECTIONAL (in a horizontal plane) and DIRECTIONAL (in a vertical plane).

From a practical viewpoint, the doublet antenna can be mounted either vertically or horizontally. The doublet shown in figure 4-14 is mounted vertically, and the radiated energy spreads out about the antenna in every direction in the horizontal plane. Since ordinarily the horizontal plane is the useful plane, this arrangement is termed NONDIRECTIONAL. The directional characteristics of the antenna in other planes is ignored. If the doublet were mounted horizontally, it would have the effect of turning the pattern on edge, reversing the patterns given in figure 4-14. The antenna would then be directional in the horizontal plane. The terms "directional" and "nondirectional" are used for convenience in describing specific radiation patterns. A complete description always involves a figure in three dimensions, as in the radiation pattern of figure 4-14.

Q17. What terms are often used to describe basic half-wave antennas?

Q18. If a basic half-wave antenna is mounted vertically, what type of radiation pattern will be produced?

Q19. In which plane will the half-wave antenna be operating if it is mounted horizontally?

RADIATION PATTERN OF A DIPOLE.—The radiation pattern of a dipole (fig. 4-15) is similar to that of the doublet (fig. 4-14). Increasing the length of the doublet to 1/2 wavelength has the effect of flattening out the radiation pattern. The radiation pattern in the horizontal plane of a dipole is a larger circle than that of the doublet. The vertical-radiation pattern lobes are no longer circular. They are flattened out and the radiation intensity is greater.

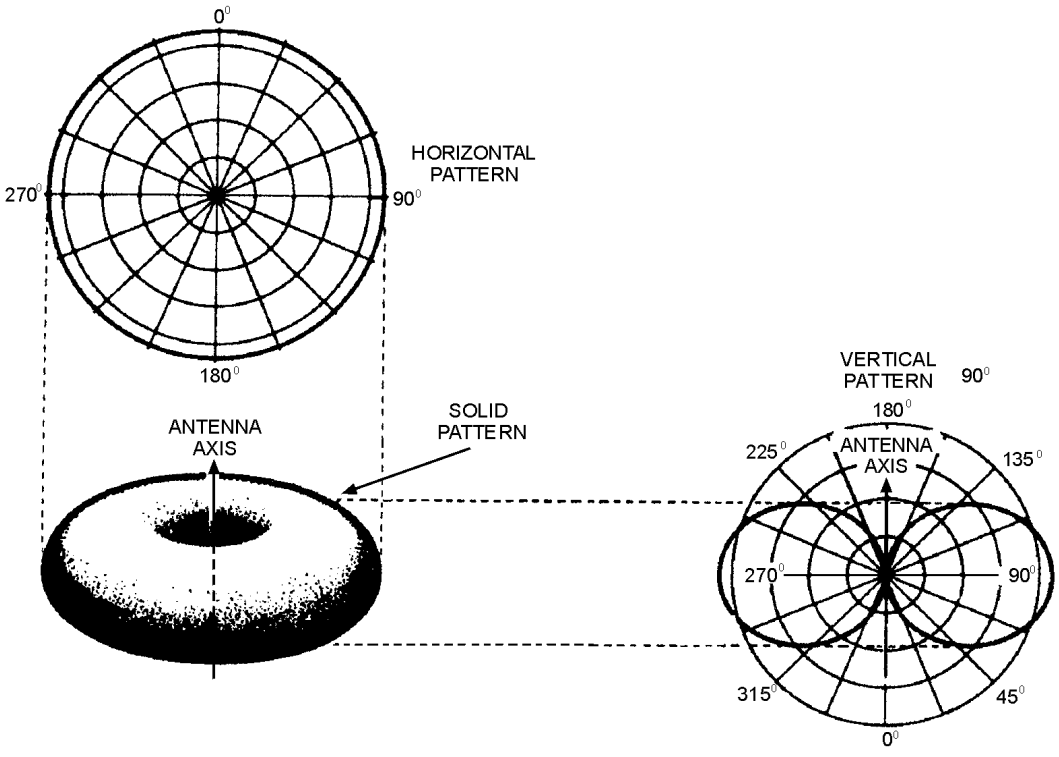


Figure 4-15.—Radiation pattern of a dipole.

Methods of Feeding Energy to an Antenna

Voltage and current distribution for the half-wave antenna (shown in figure 4-16) is the same as that for the antenna discussed earlier in this chapter. A point closely related to the voltage and current distribution on an antenna is the method of feeding the transmitter output to the antenna. The simplest method of feeding energy to the half-wave antenna is to connect one end through a capacitor to the final output stage of the transmitter. This method is often called the END-FEED or VOLTAGE-FEED method. In this method the antenna is fed at a point of high voltage (the end).

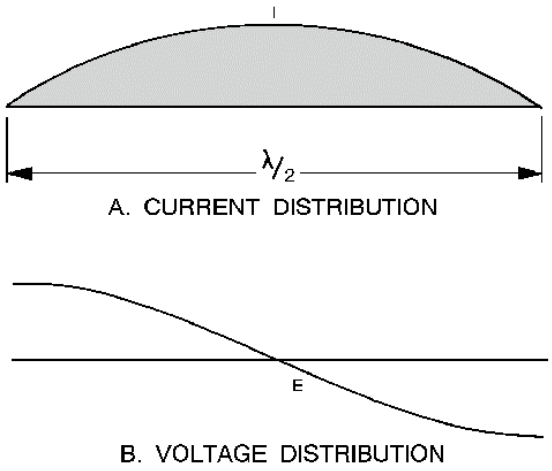


Figure 4-16.—Standing waves of current and voltage.

Energy may also be fed to the half-wave antenna by dividing the antenna at its center and connecting the transmission line from the final transmitter output stage to the two center ends of the halved antenna. Since the antenna is now being fed at the center (a point of low voltage and high current), this type of feed is known as the CENTER-FEED or CURRENT-FEED method. The point of feed is important in determining the type of transmission line to be used.

QUARTER-WAVE ANTENNAS

As you have studied in the previous sections, a 1/2 wavelength antenna is the shortest antenna that can be used in free space. If we cut a half-wave antenna in half and then ground one end, we will have a grounded quarter-wave antenna. This antenna will resonate at the same frequency as the ungrounded half-wave antenna. Such an antenna is referred to as a QUARTER-WAVE or Marconi antenna. Quarter-wave antennas are widely used in the military. Most mobile transmitting and receiving antennas (fig. 4-17) are quarter-wave antennas.

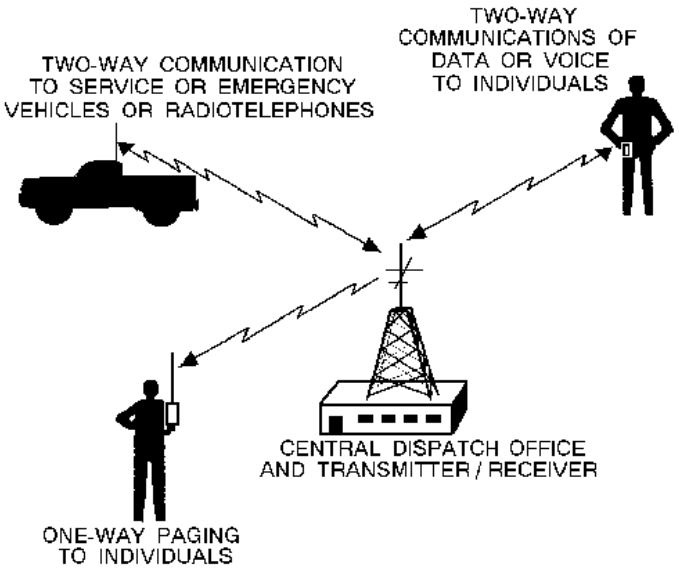


Figure 4-17.—Mobile antennas.

As stated above, a grounded quarter-wave antenna will resonate at the same frequency as an ungrounded half-wave antenna. This is because ground has high conductivity and acts as an electrical mirror image. This characteristic provides the missing half of the antenna, as shown in the bottom part of figure 4-18. In other words, the grounded quarter-wave antenna acts as if another quarter-wave were actually down in the earth.

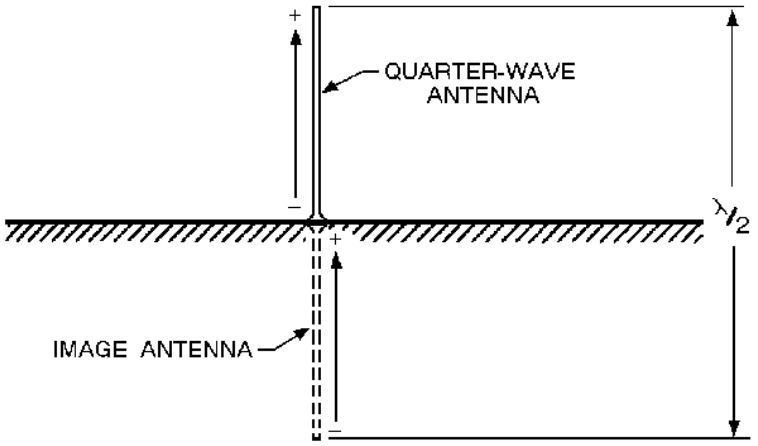


Figure 4-18.—Grounded quarter-wave antenna image.

Characteristics of Quarter-Wave Antennas

The grounded end of the quarter-wave antenna has a low input impedance and has low voltage and high current at the input end, as shown in figure 4-18. The ungrounded end has a high impedance, which causes high voltage and low current. The directional characteristics of a grounded quarter-wave antenna are the same as those of a half-wave antenna in free space.

As explained earlier, ground losses affect radiation patterns and cause high signal losses for some frequencies. Such losses may be greatly reduced if a perfectly conducting ground is provided in the

vicinity of the antenna. This is the purpose of a GROUND SCREEN (figure 4-19, view A) and COUNTERPOISE view B.

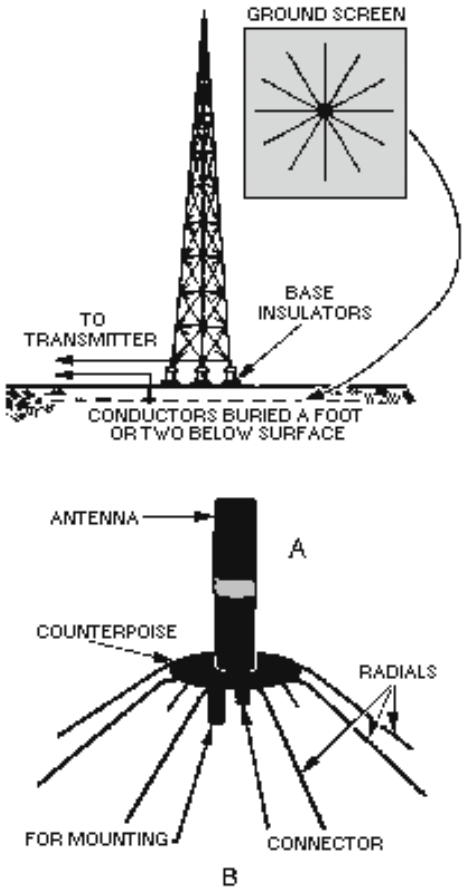


Figure 4-19.—Groundscreen and counterpoise.

The ground screen in view A is composed of a series of conductors buried 1 or 2 feet (0.3 to 0.6 meter) below the surface of the earth and arranged in a radial pattern. These conductors reduce losses in the ground in the immediate vicinity of the antenna. Such a radial system of conductors is usually 1/2 wavelength in diameter.

A counterpoise view B is used when easy access to the base of the antenna is necessary. It is also used when the earth is not a good conducting surface, such as ground that is sandy or solid rock. The counterpoise serves the same purpose as the ground screen but it is usually elevated above the earth. No specific dimensions are necessary in the construction of a counterpoise nor is the number of wires particularly critical. A practical counterpoise may be assembled from a large screen of chicken wire or some similar material. This screen may be placed on the ground, but better results are obtained if it is placed a few feet above the ground.

Q20. Since the radiation pattern of a dipole is similar to that of a doublet, what will happen to the pattern if the length of the doublet is increased?

Q21. What is the simplest method of feeding power to the half-wave antenna?

Q22. What is the radiation pattern of a quarter-wave antenna?

Q23. Describe the physical arrangement of a ground screen.

FOLDED DIPOLE

The use of parasitic elements and various stacking arrangements causes a reduction in the radiation resistance of a center-fed, half-wave antenna. Under these conditions obtaining a proper impedance match between the radiator and the transmission line is often difficult. A convenient method of overcoming these difficulties is to use a FOLDED DIPOLE in place of the center-fed radiator. (See views A and B of figure 4-20).

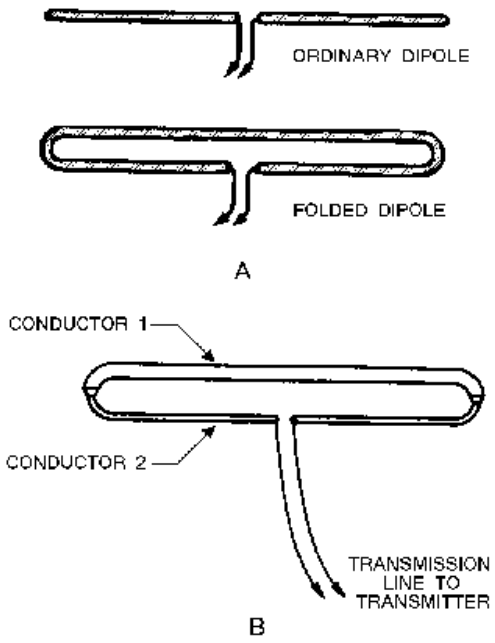


Figure 4-20.—Folded-dipole antennas.

A FOLDED DIPOLE is an ordinary half-wave antenna that has one or more additional conductors connected across its ends. Additional conductors are mounted parallel to the dipole elements at a distance equal to a very small fraction of a wavelength. Spacing of several inches is common.

The feed-point impedance can be further increased by using three or four properly spaced parallel conductors. Standard feed-line SPREADERS are used to maintain this spacing when required. In any folded dipole, the increase of impedance is the square of the number of conductors used in the radiator. Thus, a three-wire dipole has nine times (3^2) the feed-point impedance of a simple center-fed dipole. A second method of stepping up the impedance of a folded dipole is to use two conductors with different radii, as shown in view B.

The directional characteristics of a folded dipole are the same as those of a simple dipole. However, the reactance of a folded dipole varies much more slowly as the frequency is varied from resonance. Because of this the folded dipole can be used over a much wider frequency range than is possible with a simple dipole.

Q24. What is the difference in the amount of impedance between a three-wire dipole and a simple center-fed dipole?

Q25. Which has a wider frequency range, a simple dipole or a folded dipole?

ARRAY ANTENNAS

An array antenna is a special arrangement of basic antenna components involving new factors and concepts. Before you begin studying about arrays, you need to study some new terminology.

DEFINITION OF TERMS

An array antenna is made up of more than one ELEMENT, but the basic element is generally the dipole. Sometimes the basic element is made longer or shorter than a half-wave, but the deviation usually is not great.

A DRIVEN element is similar to the dipole you have been studying and is connected directly to the transmission line. It obtains its power directly from the transmitter or, as a receiving antenna, it delivers the received energy directly to the receiver. A PARASITIC ELEMENT is located near the driven element from which it gets its power. It is placed close enough to the driven element to permit coupling.

A parasitic element is sometimes placed so it will produce maximum radiation (during transmission) from its associated driver. When it operates to reinforce energy coming from the driver toward itself, the parasitic element is referred to as a DIRECTOR. If a parasitic element is placed so it causes maximum energy radiation in a direction away from itself and toward the driven element, that parasitic element is called a REFLECTOR.

If all of the elements in an array are driven, the array is referred to as a DRIVEN ARRAY (sometimes as a CONNECTED ARRAY). If one or more elements are parasitic, the entire system usually is considered to be a PARASITIC ARRAY.

MULTIELEMENT ARRAYS frequently are classified according to their directivity. A BIDIRECTIONAL ARRAY radiates in opposite directions along the line of maximum radiation. A UNIDIRECTIONAL ARRAY radiates in only one general direction.

Arrays can be described with respect to their radiation patterns and the types of elements of which they are made. However, you will find it useful to identify them by the physical placement of the elements and the direction of radiation with respect to these elements. Generally speaking, the term BROADSIDE ARRAY designates an array in which the direction of maximum radiation is perpendicular to the plane containing these elements. In actual practice, this term is confined to those arrays in which the elements themselves are also broadside, or parallel, with respect to each other.

A COLLINEAR ARRAY is one in which all the elements lie in a straight line with no radiation at the ends of the array. The direction of maximum radiation is perpendicular to the axis of the elements.

An END-FIRE ARRAY is one in which the principal direction of radiation is along the plane of the array and perpendicular to the elements. Radiation is from the end of the array, which is the reason this arrangement is referred to as an end-fire array.

Sometimes a system uses the characteristics of more than one of the three types mentioned. For instance, some of the elements may be collinear while others may be parallel. Such an arrangement is

often referred to as a COMBINATION ARRAY or an ARRAY OF ARRAYS. Since maximum radiation occurs at right angles to the plane of the array, the term broadside array is also used.

The FRONT-TO-BACK RATIO is the ratio of the energy radiated in the principal direction compared to the energy radiated in the opposite direction for a given antenna.

PHASING

Various reflected and refracted components of the propagated wave create effects of reinforcement and cancellation. At certain distant points from the transmitter, some of the wave components meet in space. Reception at these points is either impaired or improved. If the different components arrive at a given point in the same phase, they add, making a stronger signal available. If they arrive out of phase, they cancel, reducing the signal strength.

Radiation Pattern

Effects similar to those described in the preceding paragraph can be produced at the transmitting point itself. Consider the antennas shown in figure 4-21, views A and B. View A shows an unobstructed view of the radiation pattern of a single dipole. In view B two dipoles, shown as points 1 and 2, are perpendicular to the plane of the page. They are spaced 1/4 wavelength apart at the operating frequency. The radiation pattern from either antenna 1 or 2, operating alone, would be uniform in all directions in this plane, as shown in view A. Suppose that current is being fed to both antennas from the same transmitter in such a way that the current fed to antenna 2 lags the current in antenna 1 by 90 degrees. Energy radiating from antenna 1 toward receiving location X will reach antenna 2 after 1/4 cycle of operation. The energy from both antennas will add, and propagation toward X will be strong.

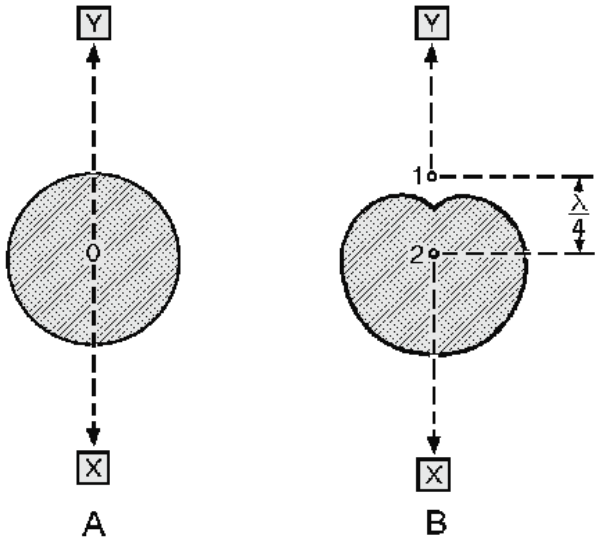


Figure 4-21.—Phasing of antenna in free space.

Radiation from antenna 2 toward receiving location Y will reach antenna 1 after 1/4 cycle. The energy in antenna 1 was 1/4 cycle behind that of antenna 2 to begin with; therefore, the radiation from antenna 1 toward receiving point Y will be exactly 180 degrees out of phase with that of antenna 2. As a result, the radiation fields will cancel and there will be no radiation toward Y.

At receiving points away from the line of radiation, phase differences occur between 0 and 180 degrees, producing varying amounts of energy in that direction. The overall effect is shown by the

radiation pattern shown in view B. The physical phase relationship caused by the 1/4-wavelength spacing between the two elements, as well as the phase of the currents in the elements, has acted to change the radiation pattern of the individual antennas.

Stub Phasing

In the case just discussed, the currents fed to the two antennas from the same transmitter were 90 degrees out of phase. Sections of transmission line, called STUBS, are frequently used for this purpose. These stubs can be adjusted to produce any desired phase relationship between connected elements.

When two collinear half-wave elements are connected directly so their currents are in the same phase, the effect is the same as that of a full-wave antenna, as shown in figure 4-22, view A. The current in the first 1/2 wavelength is exactly 180 degrees out of phase with that in the second 1/2 wavelength. This is the opposite of the desired condition. In the illustration, arrows are used to indicate the direction of current flow in the antenna. (Using arrows is a convenient means of determining the phase on more complicated arrays.)

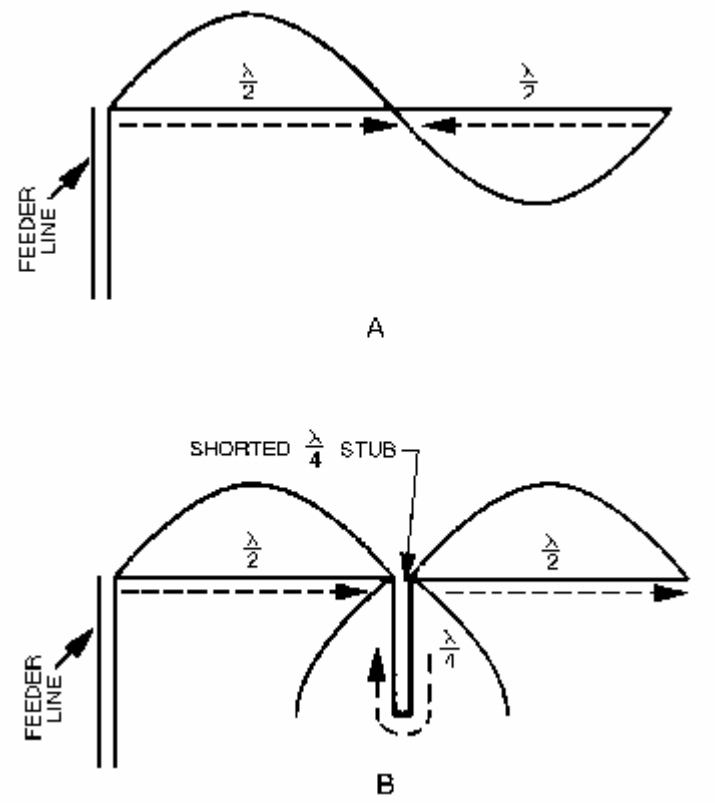


Figure 4-22.—Phasing of connected elements.

When the two elements are connected by a shorted 1/4-wavelength stub, as shown in view B, current travels down one side of the stub and up the other. It travels a distance of a 1/2 wavelength in the stub itself. As a result, the current moves through 1/2 cycle of change. When the current reaches the second element, it is in the desired phase. Since the current on one side of the stub is equal and opposite to the current on the other side, the fields produced here cancel and no radiation is transmitted from the stub itself.

DIRECTIVITY

The DIRECTIVITY of an antenna or an array can be determined by looking at its radiation pattern. In an array propagating a given amount of energy, more radiation takes place in certain directions than in others. The elements in the array can be altered in such a way that they change the pattern and distribute it more uniformly in all directions. The elements can be considered as a group of antennas fed from a common source and facing different directions. On the other hand, the elements could be arranged so that the radiation would be focused in a single direction. With no increase in power from the transmitter, the amount of radiation in a given direction would be greater. Since the input power has no increase, this increased directivity is achieved at the expense of gain in other directions.

Directivity and Interference

In many applications, sharp directivity is desirable although no need exists for added gain. Examine the physical disposition of the units shown in figure 4-23. Transmitters 1 and 2 are sending information to receivers 1 and 2, respectively, along the paths shown by the solid arrows. The distance between transmitter 1 and receiver 1 or between transmitter 2 and receiver 2 is short and does not require high-power transmission. The antennas of the transmitters propagate well in all directions. However, receiver 1 picks up some of the signals from transmitter 2, and receiver 2 picks up some of the signals from transmitter 1, as shown by the broken arrows. This effect is emphasized if the receiving antennas intercept energy equally well in all directions.

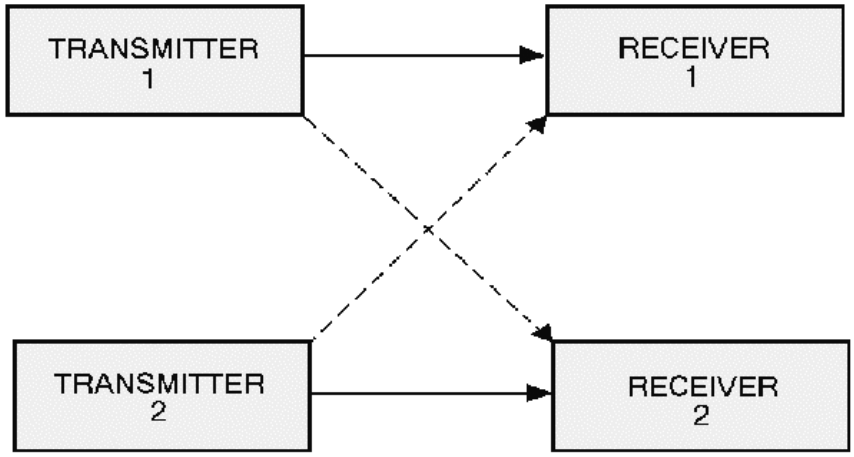


Figure 4-23.—Directivity and interference.

The use of highly directional arrays as radiators from the transmitters tends to solve the problem. The signals are beamed along the paths of the solid arrows and provide very low radiation along the paths of the broken arrows. Further improvement along these lines is obtained by the use of narrowly directed arrays as receiving antennas. The effect of this arrangement is to select the desired signal while discriminating against all other signals. This same approach can be used to overcome other types of radiated interference. In such cases, preventing radiation in certain directions is more important than producing greater gain in other directions.

Look at the differences between the field patterns of the single-element antenna and the array, as illustrated in figure 4-24. View A shows the relative field-strength pattern for a horizontally polarized single antenna. View B shows the horizontal-radiation pattern for an array. The antenna in view A

radiates fairly efficiently in the desired direction toward receiving point X. It radiates equally as efficiently toward Y, although no radiation is desired in this direction. The antenna in view B radiates strongly to point X, but very little in the direction of point Y, which results in more satisfactory operation.

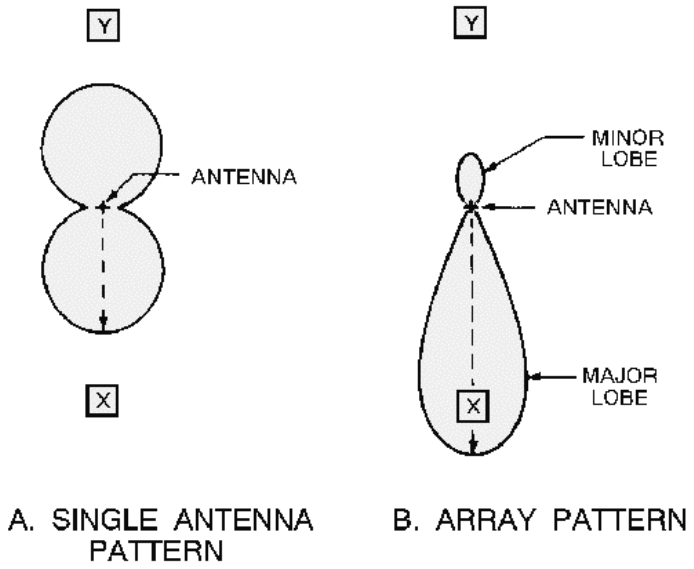


Figure 4-24.—Single antenna versus array.

Major and Minor Lobes

The pattern shown in figure 4-24, view B, has radiation concentrated in two lobes. The radiation intensity in one lobe is considerably stronger than in the other. The lobe toward point X is called a MAJOR LOBE; the other is a MINOR LOBE. Since the complex radiation patterns associated with arrays frequently contain several lobes of varying intensity, you should learn to use appropriate terminology. In general, major lobes are those in which the greatest amount of radiation occurs. Minor lobes are those in which the radiation intensity is least.

Q26. What is the purpose of antenna stubs?

Q27. What is the primary difference between the major and minor lobes of a radiation pattern?

DIRECTIONAL ARRAYS

You have already learned about radiation patterns and directivity of radiation. These topics are important to you because using an antenna with an improper radiation pattern or with the wrong directivity will decrease the overall performance of the system. In the following paragraphs, we discuss in more detail the various types of directional antenna arrays mentioned briefly in the "definition of terms" paragraph above.

Collinear Array

The pattern radiated by the collinear array is similar to that produced by a single dipole. The addition of the second radiator, however, tends to intensify the pattern. Compare the radiation pattern of the dipole (view A of figure 4-25) and the two-element antenna in view B. You will see that each pattern consists of two major lobes in opposite directions along the same axis, QQ1. There is little or no radiation along the

PP1 axis. QQ1 represents the line of maximum propagation. You can see that radiation is stronger with an added element. The pattern in view B is sharper, or more directive, than that in view A. This means that the gain along the line of maximum energy propagation is increased and the beam width is decreased. As more elements are added, the effect is heightened, as shown in view C. Unimportant minor lobes are generated as more elements are added.

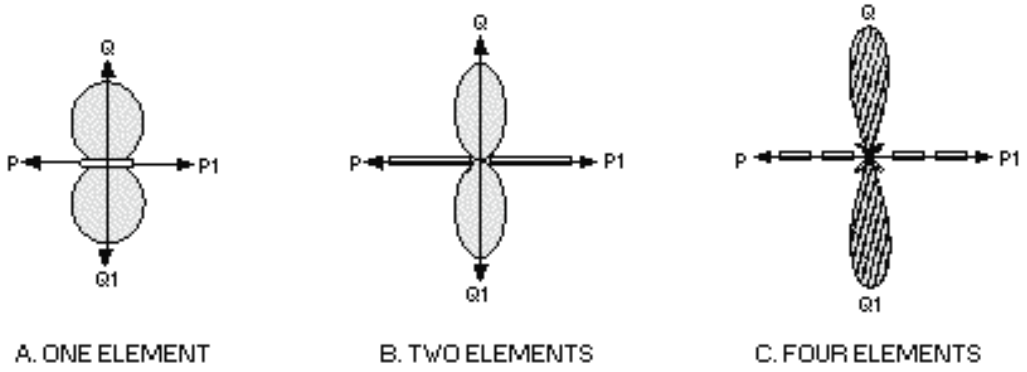


Figure 4-25.—Single half-wave antenna versus two half-wave antennas in phase.

More than four elements are seldom used because accumulated losses cause the elements farther from the point of feeding to have less current than the nearer ones. This introduces an unbalanced condition in the system and impairs its efficiency. Space limitations often are another reason for restricting the number of elements. Since this type of array is in a single line, rather than in a vertically stacked arrangement, the use of too many elements results in an antenna several wavelengths long.

RADIATION PATTERN.—The characteristic radiation pattern of a given array is obtained at the frequency or band of frequencies at which the system is resonant. The gain and directivity characteristics are lost when the antenna is not used at or near this frequency and the array tunes too sharply. A collinear antenna is more effective than an end-fire array when used off its tuned frequency. This feature is considered when transmission or reception is to be over a wide frequency band. When more than two elements are used, this advantage largely disappears.

LENGTH AND PHASING.—Although the 1/2 wavelength is the basis for the collinear element, you will find that greater lengths are often used. Effective arrays of this type have been constructed in which the elements are 0.7 and even 0.8 wavelength long. This type of array provides efficient operation at more than one frequency or over a wider frequency range. Whatever length is decided upon, all of the elements in a particular array should closely adhere to that length. If elements of different lengths are combined, current phasing and distribution are changed, throwing the system out of balance and seriously affecting the radiation pattern.

Q28. What is the maximum number of elements ordinarily used in a collinear array?

Q29. Why is the number of elements used in a collinear array limited?

Q30. How can the frequency range of a collinear array be increased?

Q31. How is directivity of a collinear array affected when the number of elements is increased?

SPACING.—The lower relative efficiency of collinear arrays of many elements, compared with other multi-element arrays, relates directly to spacing and mutual impedance effects. Mutual impedance is

an important factor to be considered when any two elements are parallel and are spaced so that considerable coupling is between them. There is very little mutual impedance between collinear sections. Where impedance does exist, it is caused by the coupling between the ends of adjacent elements. Placing the ends of elements close together is frequently necessary because of construction problems, especially where long lengths of wire are involved.

The effects of spacing and the advantages of proper spacing can be demonstrated by some practical examples. A collinear array consisting of two half-wave elements with 1/4-wavelength spacing between centers has a gain of 1.8 dB. If the ends of these same dipoles are separated so that the distance from center to center is 3/4 wavelengths and they are driven from the same source, the gain increases to approximately 2.9 dB.

A three-dipole array with negligible spacing between elements gives a gain of 3.3 dB. In other words, when two elements are used with wider spacing, the gain obtained is approximately equal to the gain obtainable from three elements with close spacing. The spacing of this array permits simpler construction, since only two dipoles are used. It also allows the antenna to occupy less space. Construction problems usually dictate small-array spacing.

Broadside Arrays

A broadside array is shown in figure 4-26, view A. Physically, it looks somewhat like a ladder. When the array and the elements in it are polarized horizontally, it looks like an upright ladder. When the array is polarized vertically, it looks like a ladder lying on one side (view B). View C is an illustration of the radiation pattern of a broadside array. Horizontally polarized arrays using more than two elements are not common. This is because the requirement that the bottom of the array be a significant distance above the earth presents construction problems. Compared with collinear arrays, broadside arrays tune sharply, but lose efficiency rapidly when not operated on the frequencies for which they are designed.

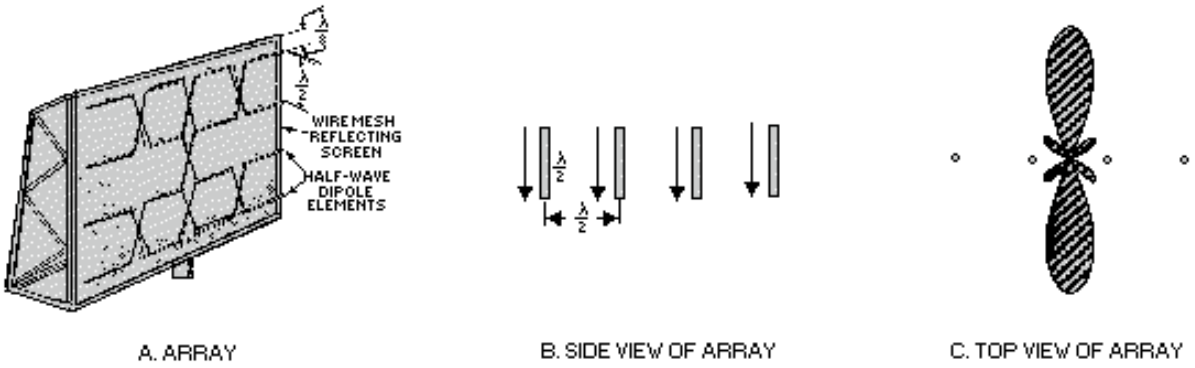


Figure 4-26.—Typical broadside array.

RADIATION PATTERN.—Figure 4-27 shows an end view of two parallel half-wave antennas (A and B) operating in the same phase and located 1/2 wavelength apart. At a point (P) far removed from the antennas, the antennas appear as a single point. Energy radiating toward P from antenna A starts out in phase with the energy radiating from antenna B in the same direction. Propagation from each antenna travels over the same distance to point P, arriving there in phase. The antennas reinforce each other in this direction, making a strong signal available at P. Field strength measured at P is greater than it would be if the total power supplied to both antennas had been fed to a single dipole. Radiation toward point P1 is built up in the same manner.

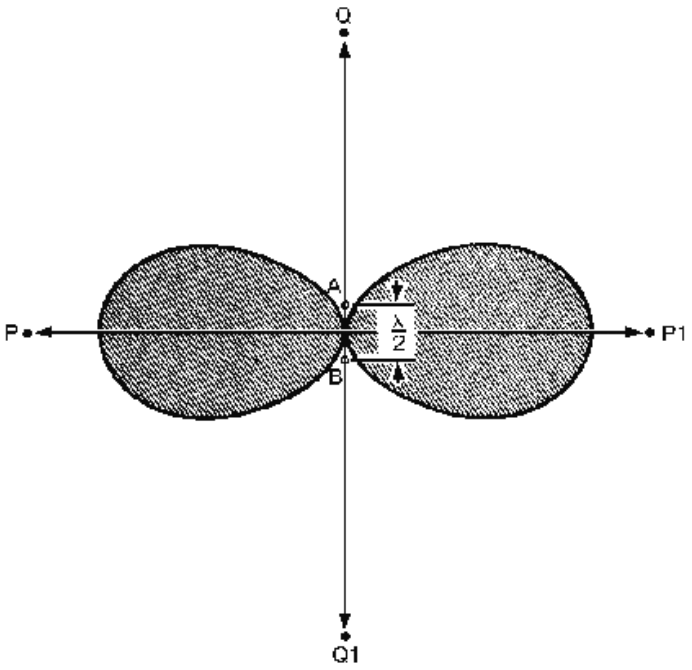


Figure 4-27.—Parallel elements in phase.

Next consider a wavefront traveling toward point Q from antenna B. By the time it reaches antenna A, 1/2 wavelength away, 1/2 cycle has elapsed. Therefore energy from antenna B meets the energy from antenna A 180 degrees out of phase. As a result, the energy moving toward point Q from the two sources cancels. In a like manner, radiation from antenna A traveling toward point Q1 meets and cancels the radiation in the same direction from antenna B. As a result, little propagation takes place in either direction along the QQ1 axis. Most of the energy is concentrated in both directions along the PP1 axis. When both antenna elements are fed from the same source, the result is the basic broadside array.

When more than two elements are used in a broadside arrangement, they are all parallel and in the same plane, as shown in figure 4-26, view B. Current phase, indicated by the arrows, must be the same for all elements. The radiation pattern shown in figure 4-26, view C, is always bi-directional. This pattern is sharper than the one shown in figure 4-27 because of the additional two elements. Directivity and gain depend on the number of elements and the spacing between them.

GAIN AND DIRECTIVITY.—The physical disposition of dipoles operated broadside to each other allows for much greater coupling between them than can occur between collinear elements. Moving the parallel antenna elements closer together or farther apart affects the actual impedance of the entire array and the overall radiation resistance as well. As the spacing between broadside elements increases, the effect on the radiation pattern is a sharpening of the major lobes. When the array consists of only two dipoles spaced exactly 1/2 wavelength apart, no minor lobes are generated at all. Increasing the distance between the elements beyond that point, however, tends to throw off the phase relationship between the original current in one element and the current induced in it by the other element. The result is that, although the major lobes are sharpened, minor lobes are introduced, even with two elements. These, however, are not large enough to be of concern.

If you add the same number of elements to both a broadside array and a collinear array, the gain of the broadside array will be greater. Reduced radiation resistance resulting from the efficient coupling between dipoles accounts for most of this gain. However, certain practical factors limit the number of

elements that may be used. The construction problem increases with the number of elements, especially when they are polarized horizontally.

Q32. What is the primary cause of broadside arrays losing efficiency when not operating at their designed frequency?

Q33. When more than two elements are used in a broadside array, how are the elements arranged?

Q34. As the spacing between elements in a broadside array increases, what is the effect on the major lobes?

End-Fire Arrays

An end-fire array looks similar to a broadside array. The ladder-like appearance is characteristic of both (fig. 4-28, view A). The currents in the elements of the end-fire array, however, are usually 180 degrees out of phase with each other as indicated by the arrows. The construction of the end-fire array is like that of a ladder lying on its side (elements horizontal). The dipoles in an end-fire array are closer together (1/8-wavelength to 1/4 -wavelength spacing) than they are for a broadside array.

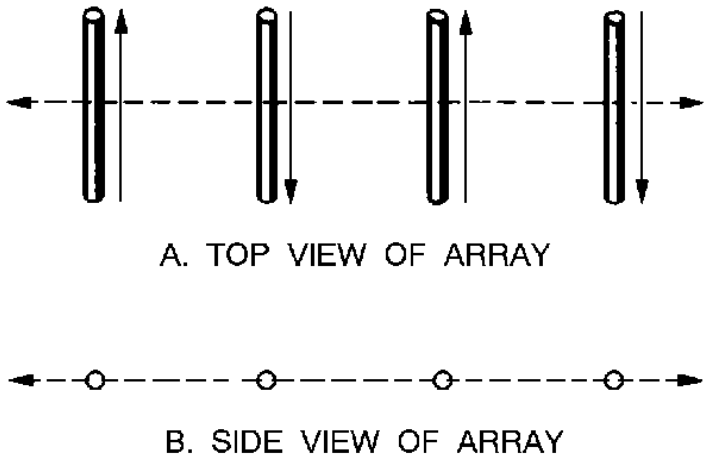


Figure 4-28.—Typical end-fire array.

Closer spacing between elements permits compactness of construction. For this reason an end-fire array is preferred to other arrays when high gain or sharp directivity is desired in a confined space. However, the close coupling creates certain disadvantages. Radiation resistance is extremely low, sometimes as low as 10 ohms, making antenna losses greater. The end-fire array is confined to a single frequency. With changes in climatic or atmospheric conditions, the danger of detuning exists.

RADIATION PATTERN.—The radiation pattern for a pair of parallel half-wave elements fed 180 degrees out of phase is shown in figure 4-29, view A. The elements shown are spaced 1/2 wavelength apart. In practice, smaller spacings are used. Radiation from elements L and M traveling toward point P begins 180 degrees out of phase. Moving the same distance over approximately parallel paths, the respective wavefronts from these elements remain 180 degrees out of phase. In other words, maximum cancellation takes place in the direction of P. The same condition is true for the opposite direction (toward P1). The P to P1 axis is the line of least radiation for the end-fire array.

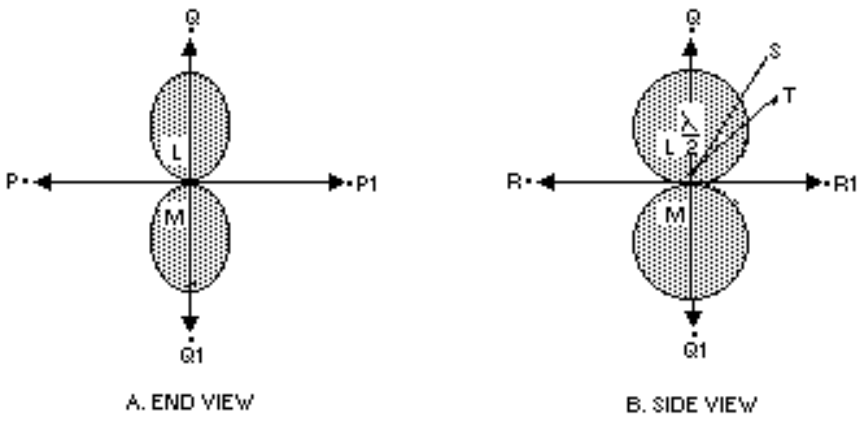


Figure 4-29.—Parallel elements 180 degrees out of phase.

Consider what happens along the QQ1 axis. Energy radiating from element M toward Q reaches element L in about 1/2 cycle (180 degrees) after it leaves its source. Since element L was fed 180 degrees out of phase with element M, the wavefronts are now in the same phase and are both moving toward Q reinforcing each other. Similar reinforcement occurs along the same axis toward Q1. This simultaneous movement towards Q and Q1 develops a bi-directional pattern. This is not always true in end-fire operation. Another application of the end-fire principle is one in which the elements are spaced 1/4 wavelength apart and phased 90 degrees from each other to produce a unidirectional pattern.

In figure 4-29, view A, elements A and B are perpendicular to the plane represented by the page; therefore, only the ends of the antennas appear. In view B the antennas are rotated a quarter of a circle in space around the QQ1 axis so that they are seen in the plane of the elements themselves. Therefore, the PP1 axis, now perpendicular to the page, is not seen as a line. The RR1 axis, now seen as a line, is perpendicular to the PP1 axis as well as to the QQ1 axis. The end-fire array is directional in this plane also, although not quite as sharply. The reason for the greater broadness of the lobes can be seen by following the path of energy radiating from the midpoint of element B toward point S in view B. This energy passes the A element at one end after traveling slightly more than the perpendicular distance between the dipoles. Energy, therefore, does not combine in exact phase toward point S. Although maximum radiation cannot take place in this direction, energy from the two sources combines closely enough in phase to produce considerable reinforcement. A similar situation exists for wavefronts traveling toward T. However, the wider angle from Q to T produces a greater phase difference and results in a decrease in the strength of the combined wave.

Directivity occurs from either one or both ends of the end-fire array, along the axis of the array, as shown by the broken arrows in figure 4-28, view A; hence, the term *end-fire* is used.

The major lobe or lobes occur along the axis of the array. The pattern is sharper in the plane that is at right angles to the plane containing the elements (figure 4-29, view A). If the elements are not exact half-wave dipoles, operation is not significantly affected. However, because of the required balance of phase relationships and critical feeding, the array must be symmetrical. Folded dipoles, such as the one shown in figure 4-20, view A, are used frequently because the impedance at their terminals is higher. This is an effective way of avoiding excessive antenna losses. Another expedient to reduce losses is the use of tubular elements of wide diameter.

GAIN AND DIRECTIVITY.—In end-fire arrays, directivity increases with the addition of more elements and with spacings approaching the optimum. The directive pattern for a two-element,

bi-directional system is illustrated in figure 4-29. View A shows radiation along the array axis in a plane perpendicular to the dipoles, and view B shows radiation along the array axis in the plane of the elements. These patterns were developed with a 180-degree phase difference between the elements. Additional elements introduce small, minor lobes.

With a 90-degree phase difference in the energy fed to a pair of end-fire elements spaced approximately 1/4 wavelength apart, unidirectional radiation can be obtained. The pattern perpendicular to the plane of the two elements is shown in figure 4-30, view A. The pattern shown in view B, taken in the same plane, is for a six-element array with 90-degree phasing between adjacent elements. Since both patterns show relative gain only, the increase in gain produced by the six-element array is not evident. End-fire arrays are the only unidirectional arrays wholly made up of driven elements.

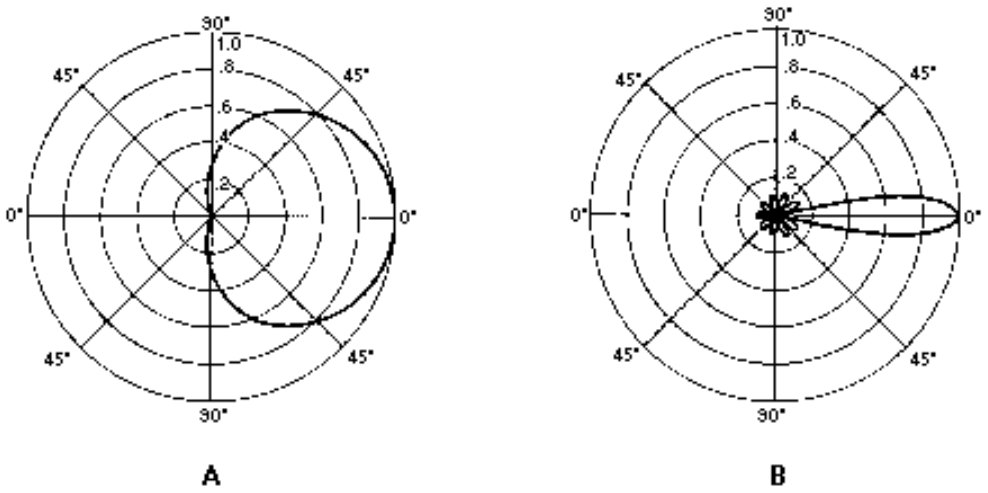


Figure 4-30.—Unidirectional end-fire arrays.

Q35. What are some disadvantages of the end-fire array?

Q36. Where does the major lobe in the end-fire array occur?

Q37. To maintain the required balance of phase relationships and critical feeding, how must the end-fire array be constructed?

Parasitic Arrays

If a small light bulb were placed in the center of a large room, the illumination would be very poor. However, if a reflector were placed behind the bulb, the space in front of the reflector would be brighter and the space behind the reflector would be dimmer. The light rays would be concentrated. Also, if a lens were placed in front of the bulb, the light would be even more concentrated and a very bright spot would appear on the wall in front of the lens. A flashlight is a practical combination of the small bulb, the reflector, and the lens. The energy from an antenna can be reflected and concentrated in a similar manner.

Although we do not usually discuss the gain of a flashlight, we can continue the comparison of an antenna and a flashlight to explain the meaning of antenna gain. Suppose the spot on the wall in front of the flashlight becomes 10 times brighter than it was when only the open bulb was used. The lens and reflector have then produced a 10-fold gain in light. For antennas, the simple half-wave antenna corresponds to the open bulb in the flashlight. Suppose an antenna system concentrates the radio waves so

that at a particular point the field strength is 10 times more than it would be at the same distance from a half-wave antenna. The antenna system is then said to have a gain of 10.

Parasitic arrays represent another method of achieving high antenna gains. A parasitic array consists of one or more parasitic elements placed in parallel with each other and, in most cases, at the same line-of-sight level. The parasitic element is fed inductively by radiated energy coming from the driven element connected to the transmitter. It is in NO way connected directly to the driven element.

When the parasitic element is placed so that it radiates away from the driven element, the element is a director. When the parasitic element is placed so that it radiates toward the driven element, the parasitic element is a reflector.

The directivity pattern resulting from the action of parasitic elements depends on two factors. These are (1) the tuning, determined by the length of the parasitic element; and (2) the spacing between the parasitic and driven elements. To a lesser degree, it also depends on the diameter of the parasitic element, since diameter has an effect on tuning.

OPERATION.—When a parasitic element is placed a fraction of a wavelength away from the driven element and is of approximately resonant length, it will re-radiate the energy it intercepts. The parasitic element is effectively a tuned circuit coupled to the driven element, much as the two windings of a transformer are coupled together. The radiated energy from the driven element causes a voltage to be developed in the parasitic element, which, in turn, sets up a magnetic field. This magnetic field extends over to the driven element, which then has a voltage induced in it. The magnitude and phase of the induced voltage depend on the length of the parasitic element and the spacing between the elements. In actual practice the length and spacing are arranged so that the phase and magnitude of the induced voltage cause a unidirectional, horizontal-radiation pattern and an increase in gain.

In the parasitic array in figure 4-31, view A, the parasitic and driven elements are spaced 1/4 wavelength apart. The radiated signal coming from the driven element strikes the parasitic element after 1/4 cycle. The voltage developed in the parasitic element is 180 degrees out of phase with that of the driven element. This is because of the distance traveled (90 degrees) and because the induced current lags the inducing flux by 90 degrees (90 + 90 = 180 degrees). The magnetic field set up by the parasitic element induces a voltage in the driven element 1/4 cycle later because the spacing between the elements is 1/4 wavelength. This induced voltage is in phase with that in the driven element and causes an increase in radiation in the direction indicated in figure 4-31, view A. Since the direction of the radiated energy is stronger in the direction away from the parasitic element (toward the driven element), the parasitic element is called a reflector. The radiation pattern as it would appear if you were looking down on the antenna is shown in view B. The pattern as it would look if viewed from the ends of the elements is shown in view C.

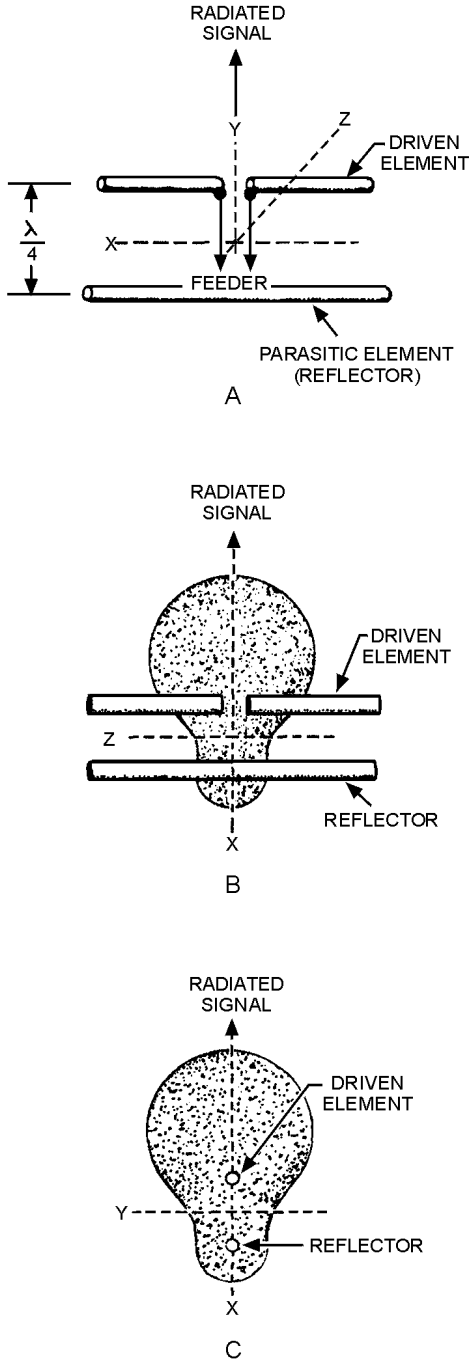


Figure 4-31.—Patterns obtained using a reflector with proper spacing.

Because the voltage induced in the reflector is 180 degrees out of phase with the signal produced at the driven element, a reduction in signal strength exists behind the reflector. Since the magnitude of an induced voltage never quite equals that of the inducing voltage, even in very closely coupled circuits, the energy behind the reflector (minor lobe) is not reduced to 0.

The spacing between the reflector and the driven element can be reduced to about 15 percent of a wavelength. The parasitic element must be made electrically inductive before it will act as a reflector. If

this element is made about 5 percent longer than 1/2 wavelength, it will act as a reflector when the spacing is 15 percent of a wavelength.

Changing the spacing and length can change the radiation pattern so that maximum radiation is on the same side of the driven element as the parasitic element. In this instance the parasitic element is called a director.

Combining a reflector and a director with the driven element causes a decrease in back radiation and an increase in directivity. This combination results in the two main advantages of a parasitic array—unidirectivity and increased gain. If the parasitic array is rotated, it can pick up or transmit in different directions because of the reduction of transmitted energy in all but the desired direction. An antenna of this type is called a ROTARY ARRAY. Size for size, both the gain and directivity of parasitic arrays are greater than those of driven arrays. The disadvantage of parasitic arrays is that their adjustment is critical and they do not operate over a wide frequency range.

GAIN AND DIRECTIVITY.—Changing the spacing between either the director or the reflector and the driven element results in a change in the radiation pattern. More gain and directivity are obtained by changing the length of the parasitic elements.

The FRONT-TO-BACK RATIO of an array is the proportion of energy radiated in the principal direction of radiation to the energy radiated in the opposite direction. A high front-to-back ratio is desirable because this means that a minimum amount of energy is radiated in the undesired direction. Since completely suppressing all such radiation is impossible, an infinite ratio cannot be achieved. In actual practice, however, rather high values can be attained. Usually the length and spacing of the parasitic elements are adjusted so that a maximum front-to-back ratio is obtained, rather than maximum gain in the desired direction.

Q38. What two factors determine the directivity pattern of the parasitic array?

Q39. What two main advantages of a parasitic array can be obtained by combining a reflector and a director with the driven element?

Q40. The parasitic array can be rotated to receive or transmit in different directions. What is the name given to such an antenna?

Q41. What are the disadvantages of the parasitic array?

Multielement Parasitic Array

A MULTIELEMENT PARASITIC array is one that contains two or more parasitic elements with the driven element. If the array contains two parasitic elements (a reflector and a director) in addition to the driven element, it is usually known as a THREE-ELEMENT ARRAY. If three parasitic elements are used, the array is known as a FOUR-ELEMENT ARRAY, and so on. Generally speaking, if more parasitic elements are added to a three-element array, each added element is a director. The field behind a reflector is so small that additional reflectors would have little effect on the overall radiation pattern. In radar, from one to five directors are used.

CONSTRUCTION.—The parasitic elements of a multi-element parasitic array usually are positioned as shown in figure 4-32, views A and B. Proper spacings and lengths are determined experimentally. A folded dipole (view B) is often used as the driven element to obtain greater values of radiation resistance.

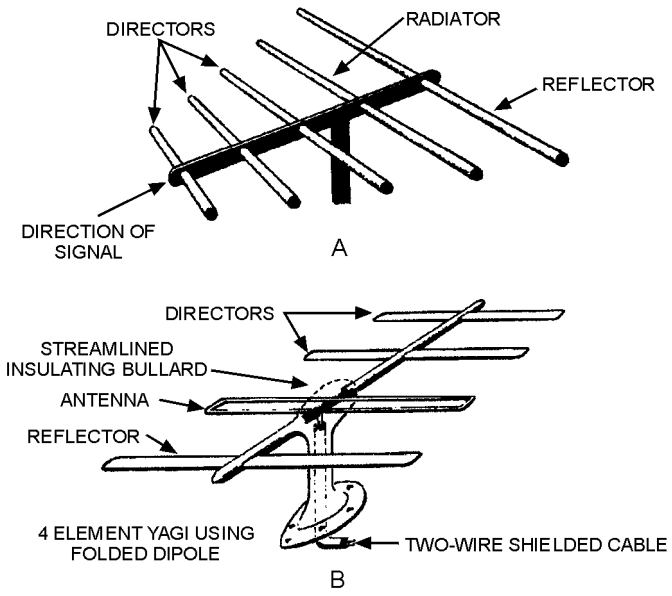


Figure 4-32.—Yagi antenna.

YAGI ANTENNAS.—An example of a multielement parasitic array is the YAGI ANTENNA (figure 4-32, views A and B). The spacings between the elements are not uniform. The radiation from the different elements arrives in phase in the forward direction, but out of phase by various amounts in the other directions.

The director and the reflector in the Yagi antenna are usually welded to a conducting rod or tube at their centers. This support does not interfere with the operation of the antenna. Since the driven element is center-fed, it is not welded to the supporting rod. The center impedance can be increased by using a folded dipole as the driven element.

The Yagi antenna shown in figure 4-32, view A, has three directors. In general, the greater number of parasitic elements used, the greater the gain. However, a greater number of such elements causes the array to have a narrower frequency response as well as a narrower beamwidth. Therefore, proper adjustment of the antenna is critical. The gain does not increase directly with the number of elements used. For example, a three-element Yagi array has a relative power gain of 5 dB. Adding another director results in a 2 dB increase. Additional directors have less and less effect.

A typical Yagi array used for receiving and transmitting energy is shown with a support frame in figure 4-33. This antenna is used by the military services. It operates at frequencies of from 12 to 50 megahertz and consists of two separate arrays (one high-frequency and one low-frequency antenna array) mounted on one frame. The various elements are indicated in the figure. The high-frequency (hf) array consists of one reflector, one driven element, and two directors; the low-frequency (lf) array has the same arrangement with one less director. The lengths of the elements in the high-frequency array are shorter than those in the low-frequency array. The physical lengths of the elements in the individual arrays are equal, but the electrical lengths can be varied by means of the tuning stubs at the center of the elements. The array can be rotated in any desired direction by a remotely controlled, electrically driven, antenna rotator.

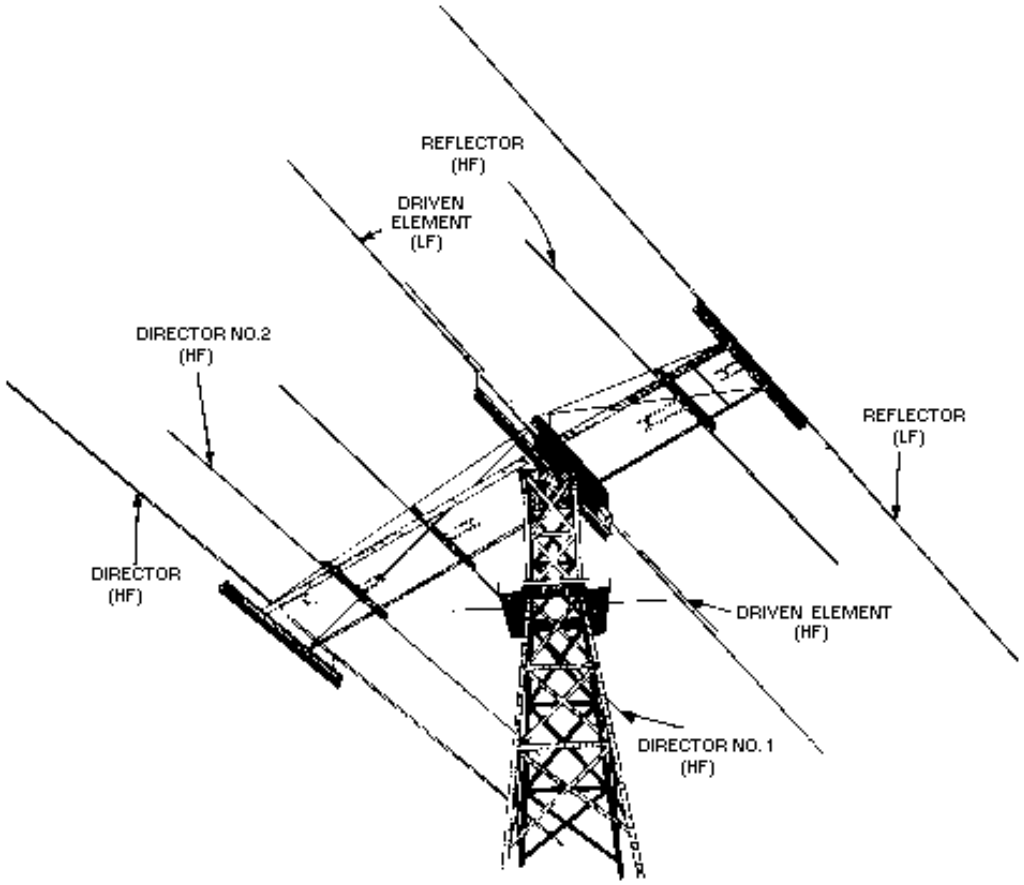


Figure 4-33.—A typical parasitic array used for transmitting and receiving.

Q42. What is the advantage of adding parasitic elements to a Yagi array?

Q43. The Yagi antenna is an example of what type of array?

SPECIAL ANTENNAS

In this section we will cover some special communications and radar antennas. Some of these antennas we touch on briefly since they are covered thoroughly in other courses.

Previously discussed antennas operate with standing waves of current and voltage along the wires. This section deals principally with antenna systems in which the current is practically uniform in all parts of the antenna. In its basic form, such an antenna consists of a single wire grounded at the far end through a resistor. The resistor has a value equal to the characteristic impedance of the antenna. This termination, just as in the case of an ordinary transmission line, eliminates standing waves. The current, therefore, decreases uniformly along the wire as the terminated end is approached. This decrease is caused by the loss of energy through radiation. The energy remaining at the end of the antenna is dissipated in the terminating resistor. For such an antenna to be a good radiator, its length must be fairly long. Also, the wire must not be too close to the ground. The return path through the ground will cause cancellation of the radiation. If the wire is sufficiently long, it will be practically nonresonant over a wide range of operating frequencies.

LONG-WIRE ANTENNA

A LONG-WIRE ANTENNA is an antenna that is a wavelength or longer at the operating frequency. In general, the gain achieved with long-wire antennas is not as great as the gain obtained from the multielement arrays studied in the previous section. But the long-wire antenna has advantages of its own. The construction of long-wire antennas is simple, both electrically and mechanically, with no particularly critical dimensions or adjustments. The long-wire antenna will work well and give satisfactory gain and directivity over a frequency range up to twice the value for which it was cut. In addition, it will accept power and radiate it efficiently on any frequency for which its overall length is not less than approximately 1/2 wavelength. Another factor is that long-wire antennas have directional patterns that are sharp in both the horizontal and vertical planes. Also, they tend to concentrate the radiation at the low vertical angles. Another type of long-wire antenna is the BEVERAGE ANTENNA, also called a WAVE ANTENNA. It is a horizontal, long-wire antenna designed especially for the reception and transmission of low-frequency, vertically polarized ground waves. It consists of a single wire, two or more wavelengths long, supported 3 to 6 meters above the ground, and terminated in its characteristic impedance, as shown in figure 4-34.

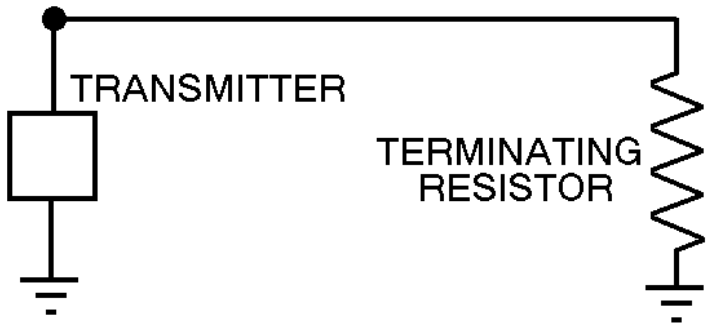


Figure 4-34.—Beverage antenna.

Q44. To radiate power efficiently, a long-wire antenna must have what minimum overall length?

Q45. What is another name for the Beverage antenna?

V ANTENNA

A V ANTENNA is a bi-directional antenna used widely in military and commercial communications. It consists of two conductors arranged to form a V. Each conductor is fed with currents of opposite polarity.

The V is formed at such an angle that the main lobes reinforce along the line bisecting the V and make a very effective directional antenna (see figure 4-35). Connecting the two-wire feed line to the apex of the V and exciting the two sides of the V 180 degrees out of phase cause the lobes to add along the line of the bisector and to cancel in other directions, as shown in figure 4-36. The lobes are designated 1, 2, 3, and 4 on leg AA', and 5, 6, 7, and 8 on leg BB'. When the proper angle between AA' and BB' is chosen, lobes 1 and 4 have the same direction and combine with lobes 7 and 6, respectively. This combination of two major lobes from each leg results in the formation of two stronger lobes, which lie along an imaginary line bisecting the enclosed angle. Lobes 2, 3, 5, and 8 tend to cancel each other, as do the smaller lobes, which are approximately at right angles to the wire legs of the V. The resultant waveform pattern is shown at the right of the V antenna in figure 4-36.

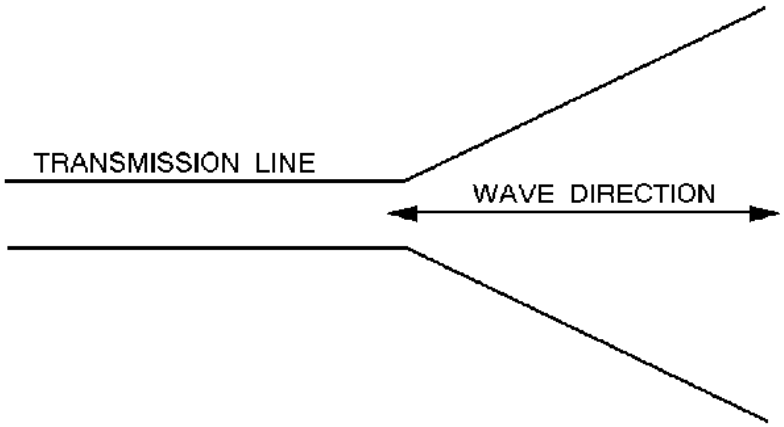


Figure 4-35.—Basic V antenna.

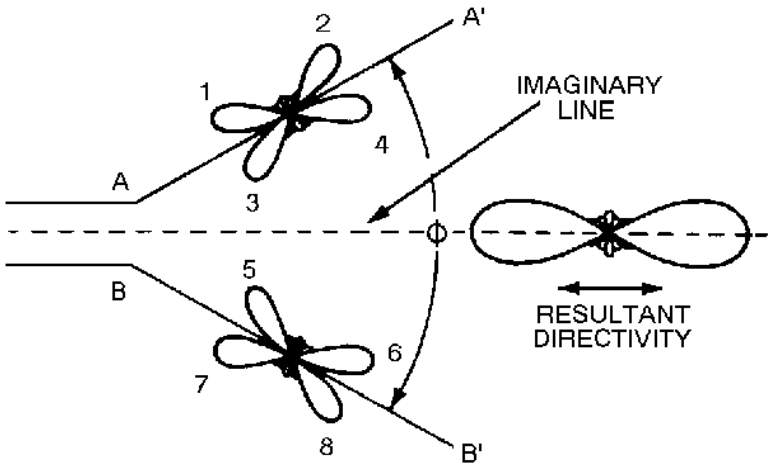


Figure 4-36.—Formation of directional radiation pattern from a resonant V antenna.

Q46. What is the polarity of the currents that feed the V antenna?

RHOMBIC ANTENNA

The highest development of the long-wire antenna is the RHOMBIC ANTENNA (see figure 4-37). It consists of four conductors joined to form a rhombus, or diamond shape. The antenna is placed end to end and terminated by a noninductive resistor to produce a uni-directional pattern. A rhombic antenna can be made of two obtuse-angle V antennas that are placed side by side, erected in a horizontal plane, and terminated so the antenna is nonresonant and unidirectional.

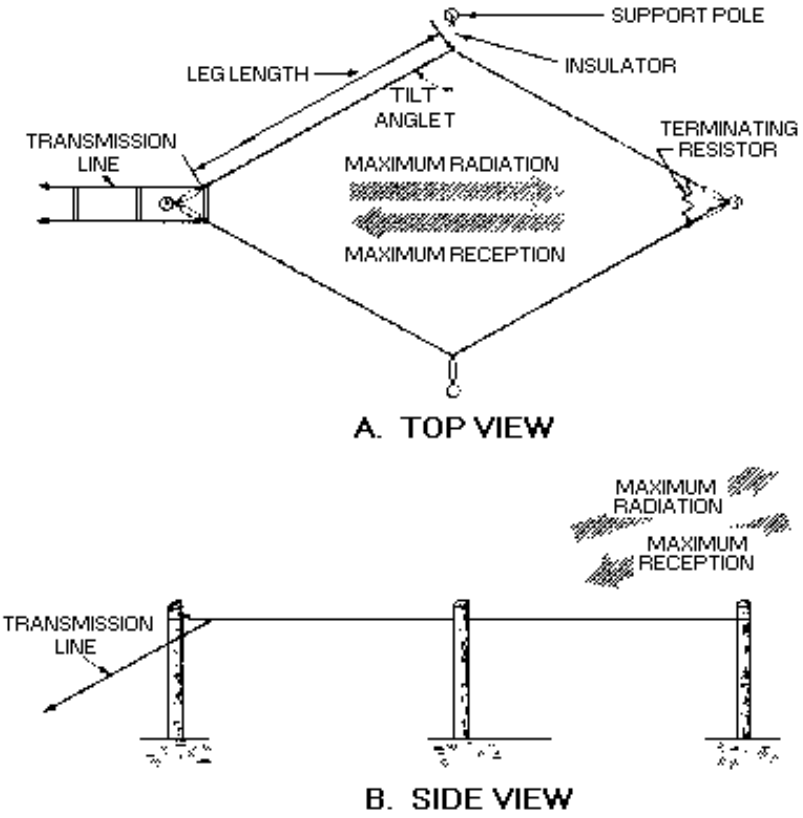


Figure 4-37.—Basic rhombic antenna.

The rhombic antenna is WIDELY used for long-distance, high-frequency transmission and reception. It is one of the most popular fixed-station antennas because it is very useful in point-to-point communications.

Advantages

The rhombic antenna is useful over a wide frequency range. Although some changes in gain, directivity, and characteristic impedance do occur with a change in operating frequency, these changes are small enough to be neglected.

The rhombic antenna is much easier to construct and maintain than other antennas of comparable gain and directivity. Only four supporting poles of common heights from 15 to 20 meters are needed for the antenna.

The rhombic antenna also has the advantage of being noncritical as far as operation and adjustment are concerned. This is because of the broad frequency characteristics of the antenna.

Still another advantage is that the voltages present on the antenna are much lower than those produced by the same input power on a resonant antenna. This is particularly important when high transmitter powers are used or when high-altitude operation is required.

Disadvantages

The rhombic antenna is not without its disadvantages. The principal one is that a fairly large antenna site is required for its erection. Each leg is made at least 1 or 2 wavelengths long at the lowest operating frequency. When increased gain and directivity are required, legs of from 8 to 12 wavelengths are used. These requirements mean that high-frequency rhombic antennas have wires of several hundred feet in length. Therefore, they are used only when a large plot of land is available.

Another disadvantage is that the horizontal and vertical patterns depend on each other. If a rhombic antenna is made to have a narrow horizontal beam, the beam is also lower in the vertical direction. Therefore, obtaining high vertical-angle radiation is impossible except with a very broad horizontal pattern and low gain. Rhombic antennas are used, however, for long-distance sky wave coverage at the high frequencies. Under these conditions low vertical angles of radiation (less than 20 degrees) are desirable. With the rhombic antenna, a considerable amount of the input power is dissipated uselessly in the terminating resistor. However, this resistor is necessary to make the antenna unidirectional. The great gain of the antenna more than makes up for this loss.

Radiation Patterns

Figure 4-38 shows the individual radiation patterns produced by the four legs of the rhombic antenna and the resultant radiation pattern. The principle of operation is the same as for the V and the half-rhombic antennas.

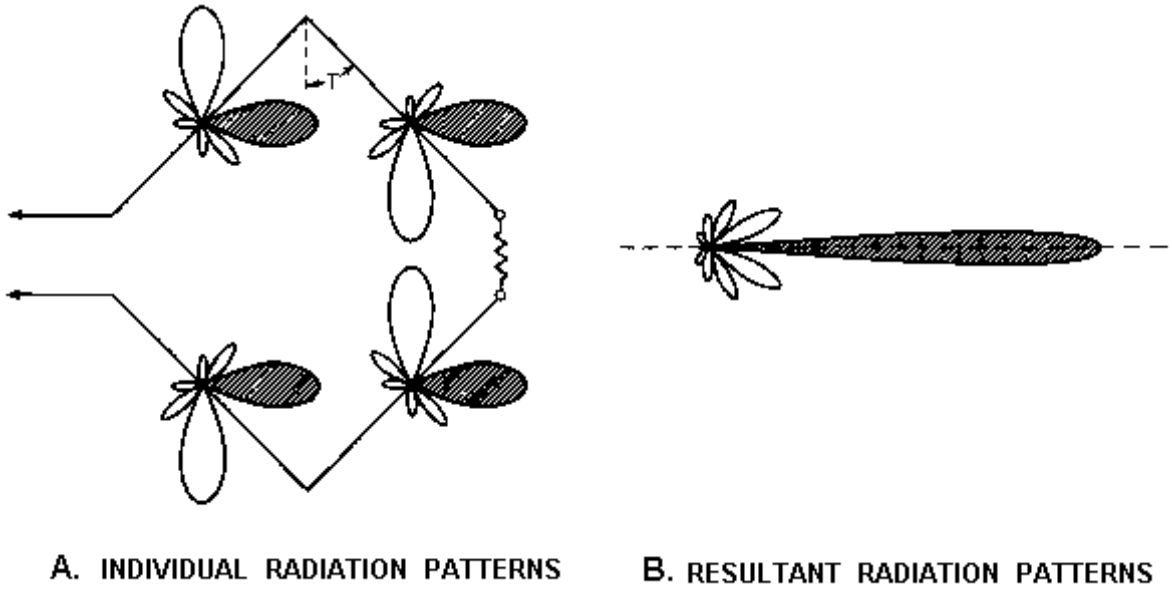


Figure 4-38.—Formation of a rhombic antenna beam.

Terminating Resistor

The terminating resistor plays an important part in the operation of the rhombic antenna. Upon it depend the unidirectivity of the antenna and the lack of resonance effects. An antenna should be properly terminated so it will have a constant impedance at its input. Terminating the antenna properly will also allow it to be operated over a wide frequency range without the necessity for changing the coupling adjustments at the transmitter. Discrimination against signals coming from the rear is of great importance

for reception. The reduction of back radiation is perhaps of lesser importance for transmission. When an antenna is terminated with resistance, the energy that would be radiated backward is absorbed in the resistor.

Q47. What is the main disadvantage of the rhombic antenna?

TURNSTILE ANTENNA

The TURNSTILE ANTENNA is one of the many types that has been developed primarily for omnidirectional vhf communications. The basic turnstile consists of two horizontal half-wave antennas mounted at right angles to each other in the same horizontal plane. When these two antennas are excited with equal currents 90 degrees out of phase, the typical figure-eight patterns of the two antennas merge to produce the nearly circular pattern shown in figure 4-39, view A. Pairs of such antennas are frequently stacked, as shown in figure 4-40. Each pair is called a BAY. In figure 4-40 two bays are used and are spaced 1/2 wavelength apart, and the corresponding elements are excited in phase. These conditions cause a part of the vertical radiation from each bay to cancel that of the other bay. This results in a decrease in energy radiated at high vertical angles and increases the energy radiated in the horizontal plane. Stacking a number of bays can alter the vertical radiation pattern, causing a substantial gain in a horizontal direction without altering the overall horizontal directivity pattern. Figure 4-39, view B, compares the circular vertical radiation pattern of a single-bay turnstile with the sharp pattern of a four-bay turnstile array. A three-dimensional radiation pattern of a four-bay turnstile antenna is shown in figure 4-39, view C.

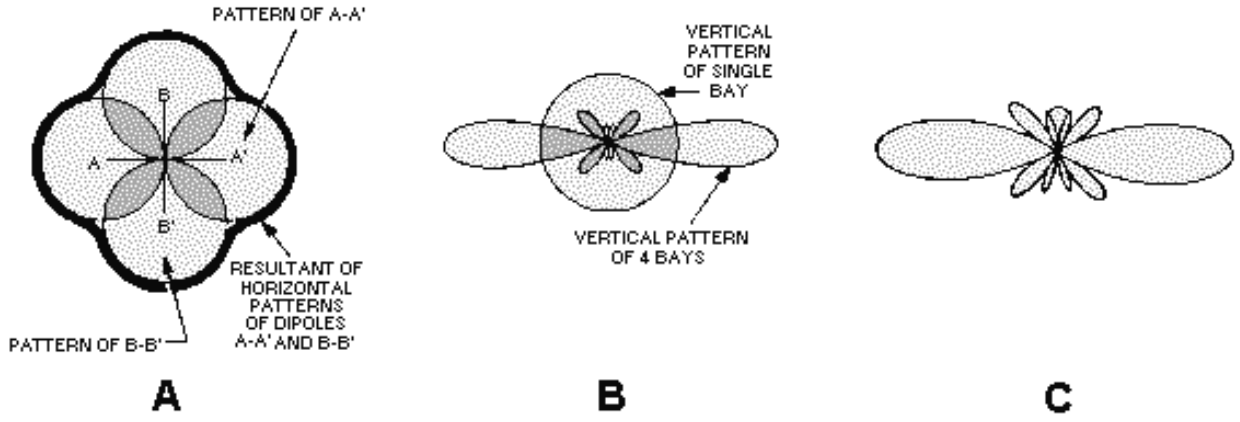


Figure 4-39.—Turnstile antenna radiation pattern.

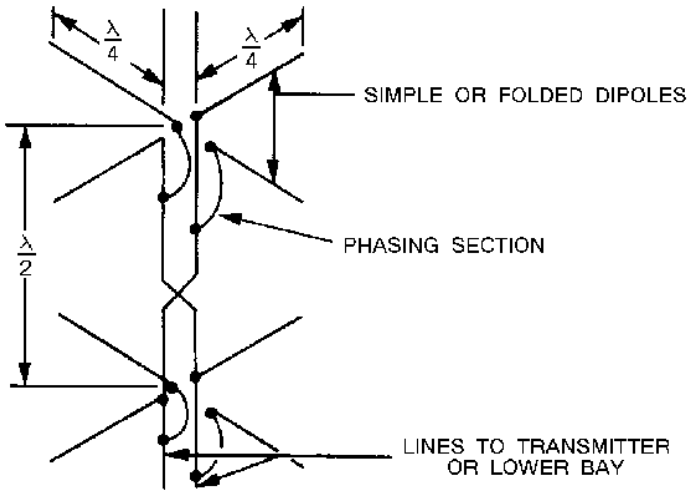


Figure 4-40.—Stacked turnstile antennas.

GROUND-PLANE ANTENNA

A vertical quarter-wave antenna several wavelengths above ground produces a high angle of radiation that is very undesirable at vhf and uhf frequencies. The most common means of producing a low angle of radiation from such an antenna is to work the radiator against a simulated ground called a GROUND PLANE. A simulated ground may be made from a large metal sheet or several wires or rods radiating from the base of the radiator. An antenna so constructed is known as a GROUND-PLANE ANTENNA. Two ground-plane antennas are shown in figure 4-41, views A and B.

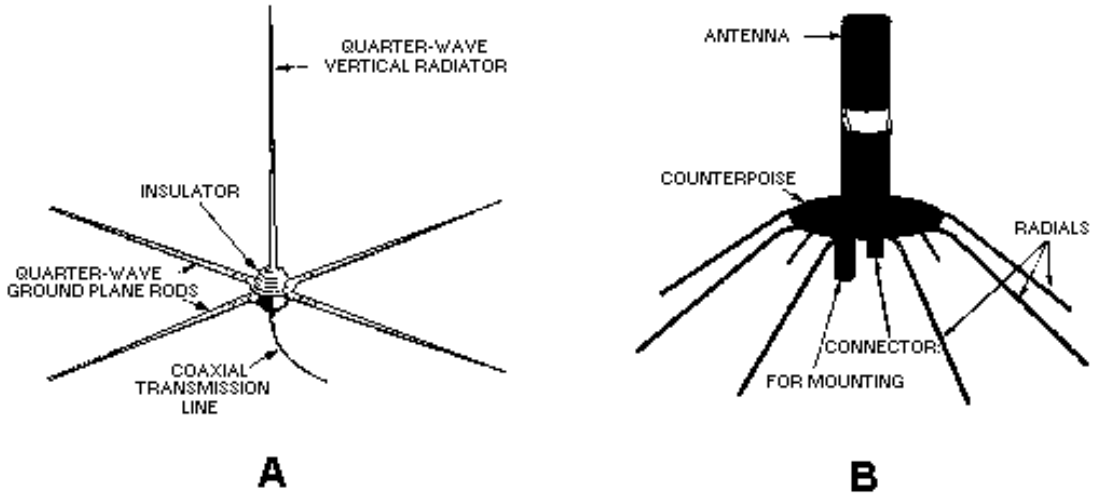


Figure 4-41.—Ground-plane antennas.

CORNER REFLECTOR

When a unidirectional radiation pattern is desired, it can be obtained by the use of a corner reflector with a half-wave dipole. A CORNER-REFLECTOR ANTENNA is a half-wave radiator with a reflector. The reflector consists of two flat metal surfaces meeting at an angle immediately behind the radiator. In other words, the radiator is set in the plane of a line bisecting the corner angle formed by the reflector

sheets. The construction of a corner reflector is shown in figure 4-42. Corner-reflector antennas are mounted with the radiator and the reflector in the horizontal position when horizontal polarization is desired. In such cases the radiation pattern is very narrow in the vertical plane, with maximum signal being radiated in line with the bisector of the corner angle. The directivity in the horizontal plane is approximately the same as for any half-wave radiator having a single-rod type reflector behind it. If the antenna is mounted with the radiator and the corner reflector in the vertical position, as shown in view A, maximum radiation is produced in a very narrow horizontal beam. Radiation in a vertical plane will be the same as for a similar radiator with a single-rod type reflector behind it.

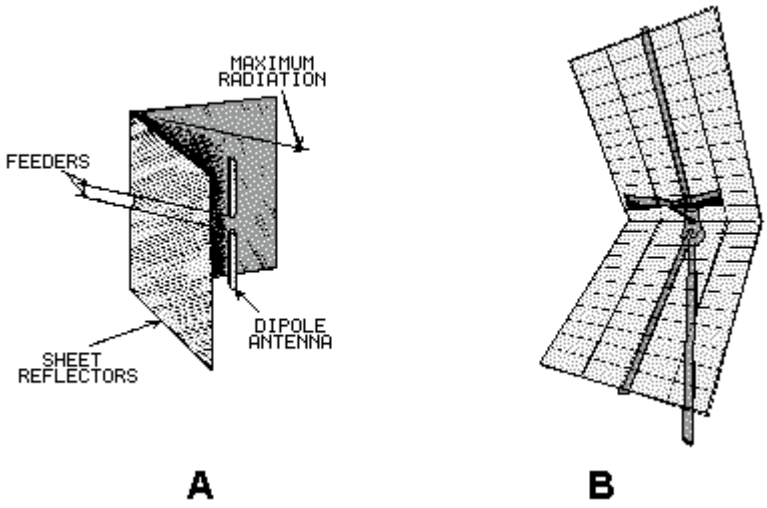


Figure 4-42.—Corner-reflector antennas.

Q48. What is the primary reason for the development of the turnstile antenna?

RF SAFETY PRECAUTIONS

Although electromagnetic radiation from transmission lines and antennas is usually of insufficient strength to electrocute personnel, it can lead to other accidents and compound injuries. Voltages may be induced in ungrounded metal objects, such as wire guys, wire cable (hawser), hand rails, or ladders. If you come in contact with these objects, you could receive a shock or rf burn. This shock can cause you to jump or fall into nearby mechanical equipment or, when working aloft, to fall from an elevated work area. Take care to ensure that all transmission lines or antennas are deenergized before working near or on them.

Either check or have someone check all guys, cables, rails, and ladders around your work area for rf shock dangers. Use working aloft "chits" and safety harnesses for your own safety. Signing a "working aloft chit" signifies that all equipment is in a nonradiating status. The person who signs the chit should ensure that no rf danger exists in areas where you or other personnel will be working.

Nearby ships or parked aircraft are another source of rf energy that you must consider when you check a work area for safety. Combustible materials can be ignited and cause severe fires from arcs or heat generated by rf energy. Also, rf radiation can detonate ordnance devices by inducing currents in the internal wiring of the devices or in the external test equipment or leads connected to them.

ALWAYS obey rf radiation warning signs and keep a safe distance from radiating antennas. The six types of warning signs for rf radiation hazards are shown in figure 4-43.

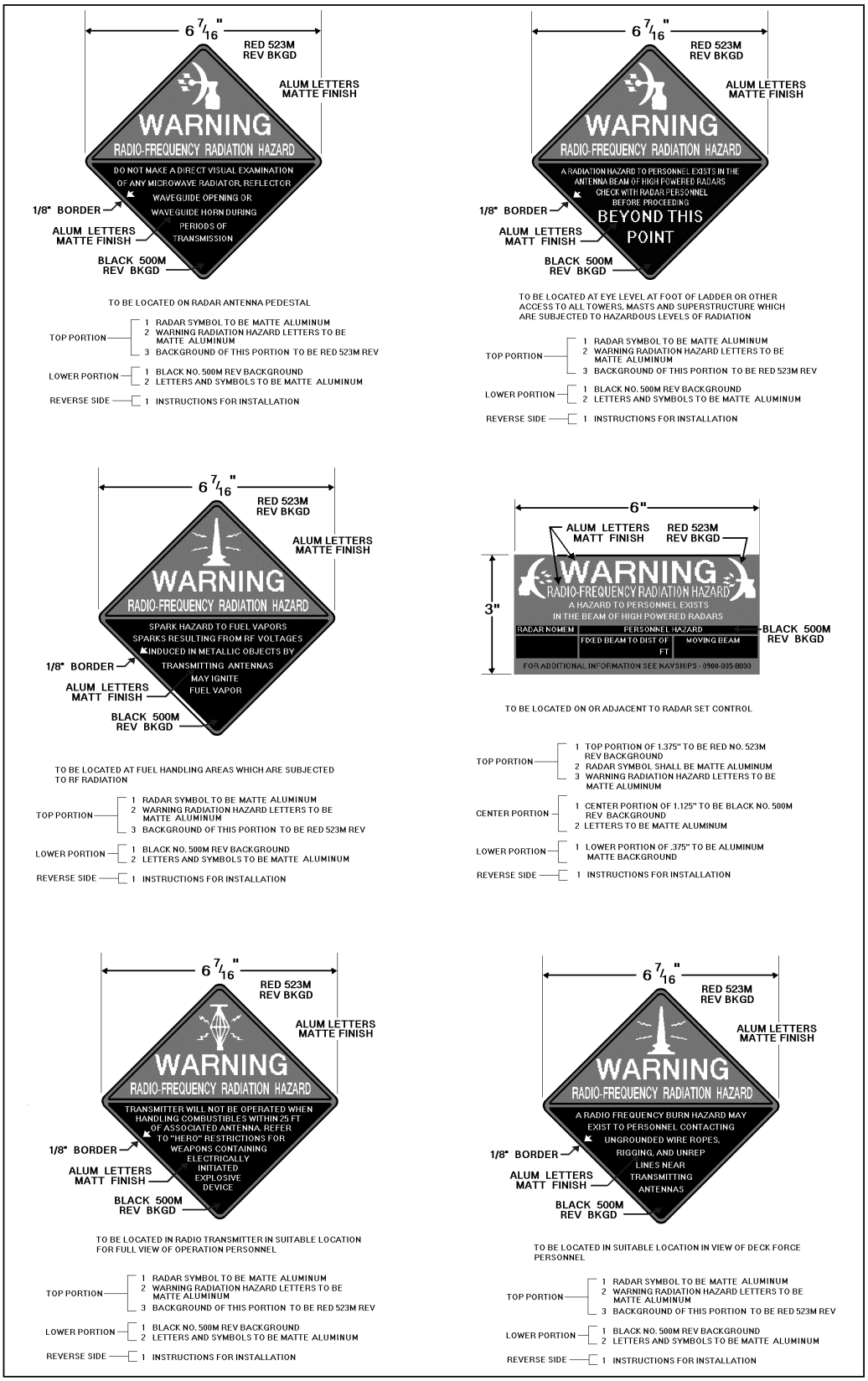


Figure 4-43.—Examples of rf radiation warning signs.

RF BURNS

Close or direct contact with rf transmission lines or antennas may result in rf burns. These are usually deep, penetrating, third-degree burns. To heal properly, these burns must heal from the inside to the skin's surface. To prevent infection, you must give proper attention to all rf burns, including the small "pinhole" burns. Petrolatum gauze can be used to cover these burns temporarily, before the injured person reports to medical facilities for further treatment.

DIELECTRIC HEATING

DIELECTRIC HEATING is the heating of an insulating material by placing it in a high-frequency electric field. The heat results from internal losses during the rapid reversal of polarization of molecules in the dielectric material.

In the case of a human in an rf field, the body acts as a dielectric. If the power in the rf field exceeds 10 milliwatts per centimeter, a person in that field will have a noticeable rise in body temperature. The eyes are highly susceptible to dielectric heating. For this reason, you should not look directly into devices radiating rf energy. The vital organs of the body also are susceptible to dielectric heating. For your own safety, you must NOT stand directly in the path of rf radiating devices.

PRECAUTIONS WHEN WORKING ALOFT

When radio or radar antennas are energized by transmitters, you must not go aloft unless advance tests show that little or no danger exists. A casualty can occur from even a small spark drawn from a charged piece of metal or rigging. Although the spark itself may be harmless, the "surprise" may cause you to let go of the antenna involuntarily and you may fall. There is also a shock hazard if nearby antennas are energized.

Rotating antennas also might cause you to fall when you are working aloft. Motor safety switches controlling the motion of rotating antennas must be tagged and locked open before you go aloft near such antennas.

When working near a stack, you should draw and wear the recommended oxygen breathing apparatus. Among other toxic substances, stack gas contains carbon monoxide. Carbon monoxide is too unstable to build up to a high concentration in the open, but prolonged exposure to even small quantities is dangerous.

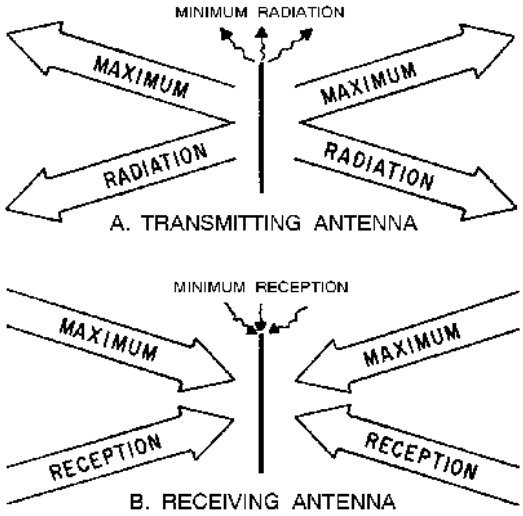
SUMMARY

This chapter has presented information on the various types of antennas. The information that follows summarizes the important points of this chapter.

An **ANTENNA** is a conductor, or system of conductors, that radiates or receives energy in the form of electromagnetic waves.

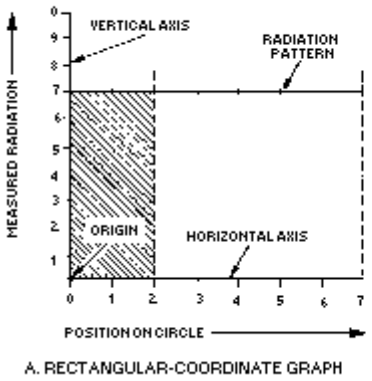
HERTZ (half-wave) and **MARCONI** (quarter-wave) are the two basic classifications of antennas.

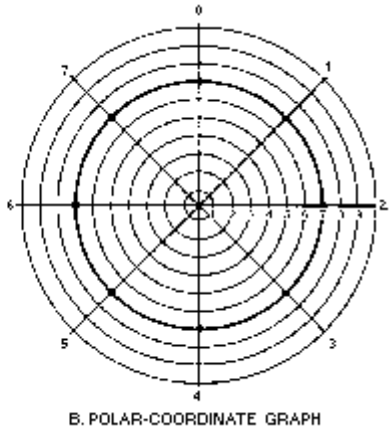
RECIPROCITY of antennas means that the various properties of the antenna apply equally to transmitting and receiving.



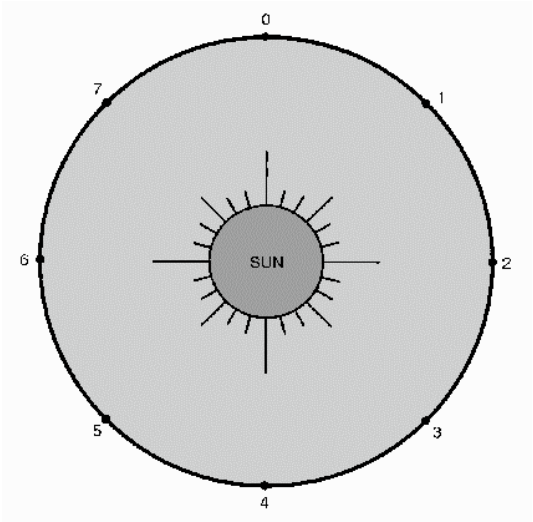
RADIATION RESISTANCE is the amount of resistance which, if inserted in place of the antenna, would consume the same amount of power that is actually radiated by the antenna.

RADIATION PATTERNS can be plotted on a rectangular- or polar-coordinate graph. These patterns are a measurement of the energy leaving an antenna.

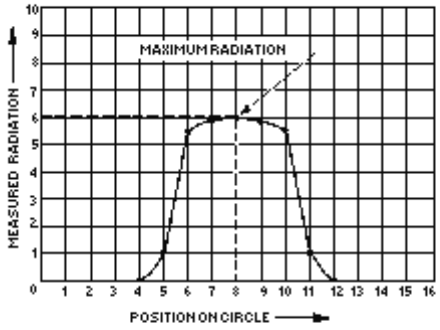




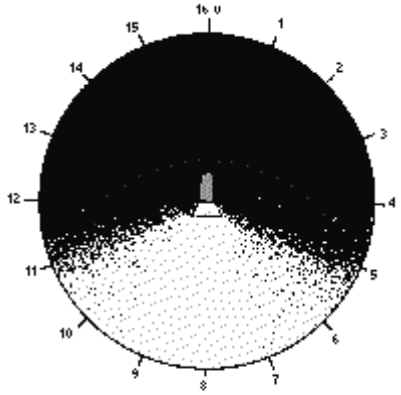
An **ISOTROPIC RADIATOR** radiates energy equally in all directions.



An **ANISOTROPIC RADIATOR** radiates energy directionally.



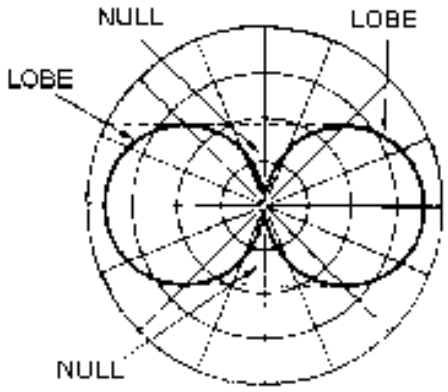
A



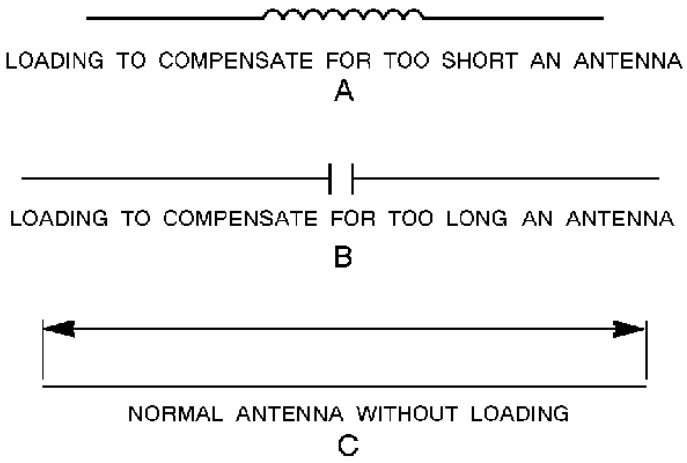
B

A **LOBE** is the area of a radiation pattern that is covered by radiation.

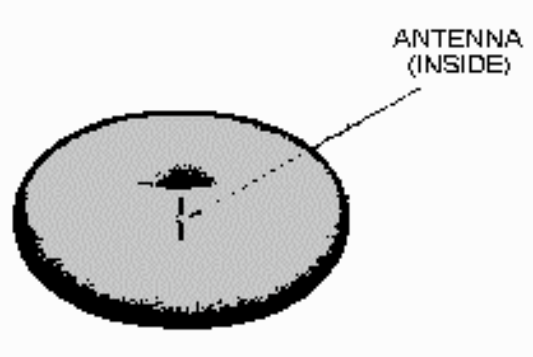
A **NULL** is the area of a radiation pattern that has minimum radiation.



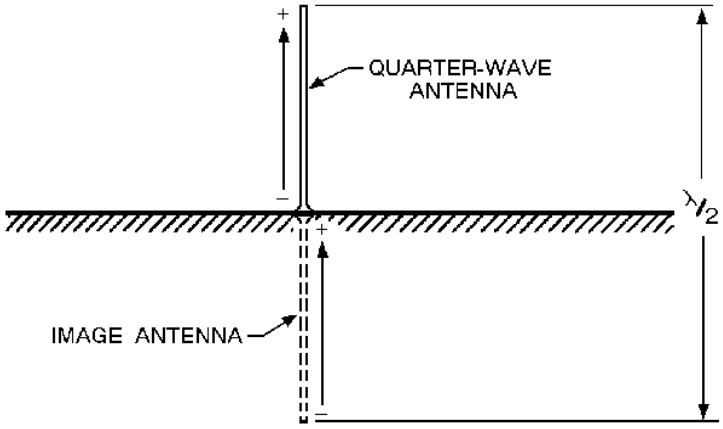
ANTENNA LOADING is the method used to change the electrical length of an antenna. This keeps the antenna in resonance with the applied frequency. It is accomplished by inserting a variable inductor or capacitor in series with the antenna.



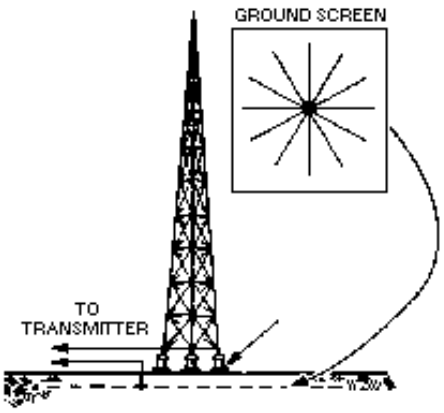
A **HALF-WAVE ANTENNA (Hertz)** consists of two lengths of rod or tubing, each a quarter-wave long at a certain frequency, which radiates a doughnut pattern.

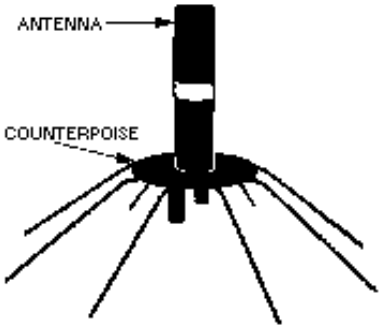


A **QUARTER-WAVE ANTENNA (Marconi)** is a half-wave antenna cut in half with one end grounded. The ground furnishes the missing half of the antenna.

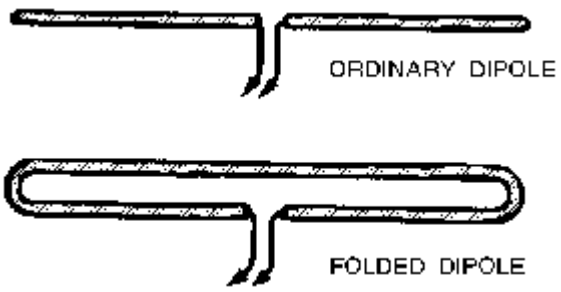


The **GROUND SCREEN** and the **COUNTERPOISE** are used to reduce losses caused by the ground in the immediate vicinity of the antenna. The ground screen is buried below the surface of the earth. The counterpoise is installed above the ground.





The **FOLDED DIPOLE** consists of a dipole radiator, which is connected in parallel at its ends to a half-wave radiator.



AN **ARRAY** is a combination of half-wave elements operating together as a single antenna. It provides more gain and greater directivity than single element antennas.

A **DRIVEN ARRAY** derives its power directly from the source.

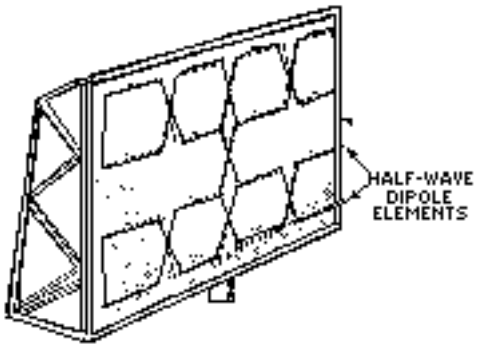
A **PARASITIC ARRAY** derives its power by coupling the energy from other elements of the antenna.

The **BIDIRECTIONAL ARRAY** radiates energy equally in two opposing directions.

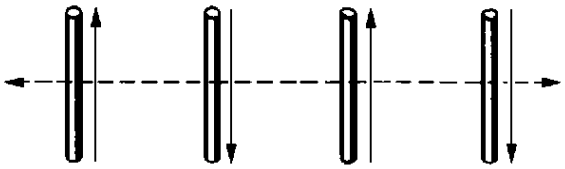
The **UNIDIRECTIONAL ARRAY** radiates energy efficiently in a single direction.

The **COLLINEAR ARRAY** has elements in a straight line. Maximum radiation occurs at right angles to this line.

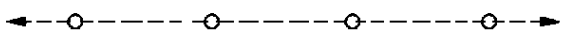
The **BROADSIDE ARRAY** has elements parallel and in the same plane. Maximum radiation develops in the plane at right angles to the plane of the elements.



The **END-FIRE ARRAY** has elements parallel to each other and in the same plane. Maximum radiation occurs along the axis of the array.



A. TOP VIEW OF ARRAY



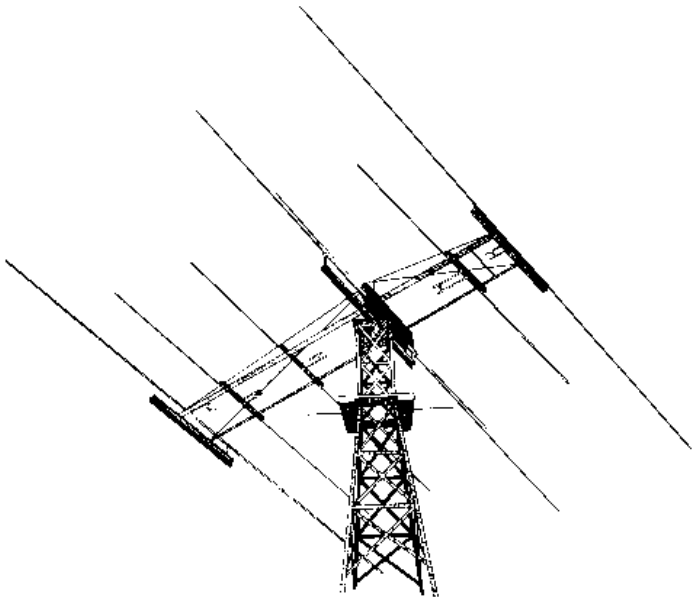
B. SIDE VIEW OF ARRAY

MATCHING STUBS are used between elements to maintain current in the proper phase.

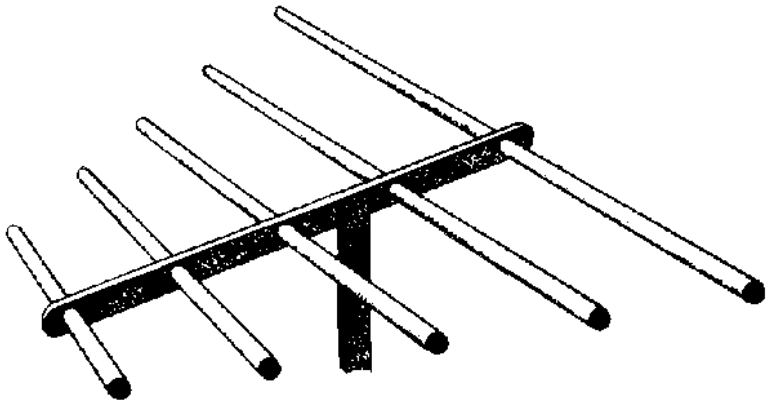
The **GAIN OF A COLLINEAR ANTENNA** is greatest when the elements are spaced from 0.4 to 0.5 wavelength apart or when the number of elements is increased.

The **OPTIMUM GAIN OF A BROADSIDE ARRAY** is obtained when the elements are spaced 0.65 wavelength apart.

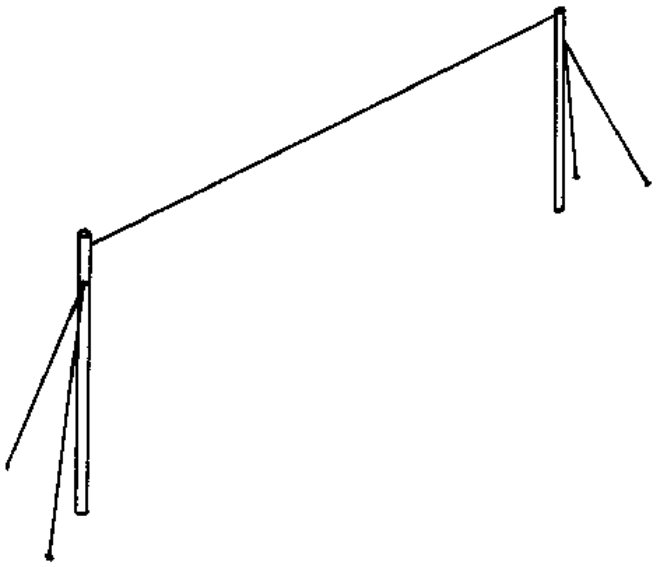
A **PARASITIC ARRAY** consists of one or more parasitic elements with a driven element. The amount of power gain and directivity depends on the lengths of the parasitic elements and the spacing between them.



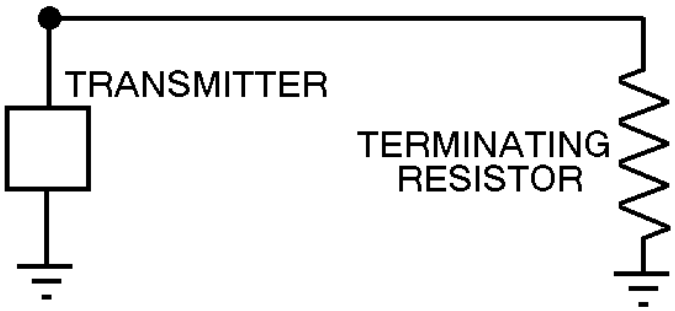
MULTIELEMENT ARRAYS, such as the YAGI, have a narrow frequency response as well as a narrow beamwidth.



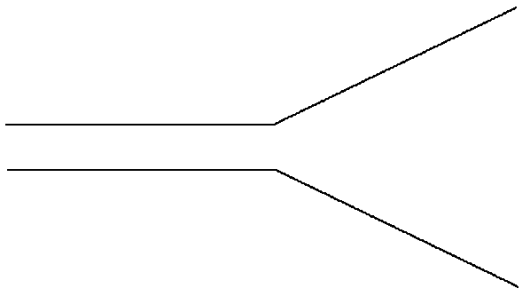
A **LONG-WIRE ANTENNA** is an antenna that is a wavelength or more long at the operating frequency. These antennas have directive patterns that are sharp in both the horizontal and vertical planes.



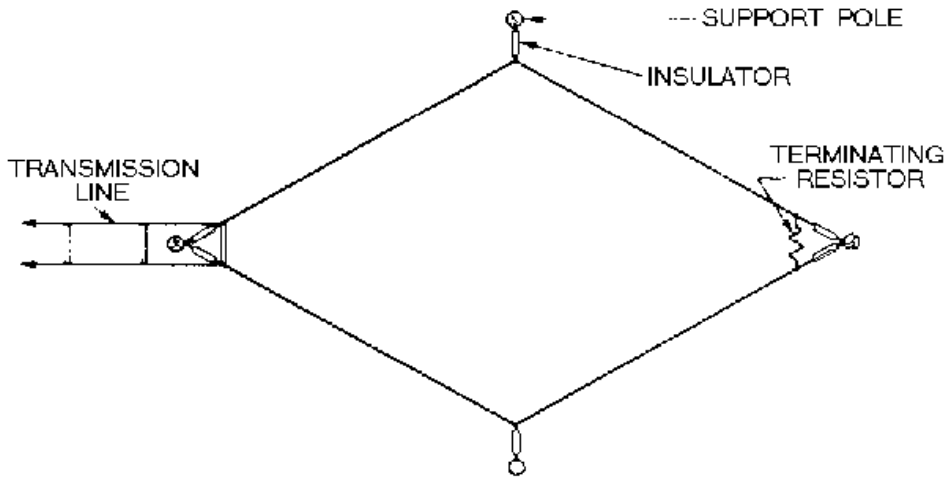
BEVERAGE ANTENNAS consist of a single wire that is two or more wavelengths long.



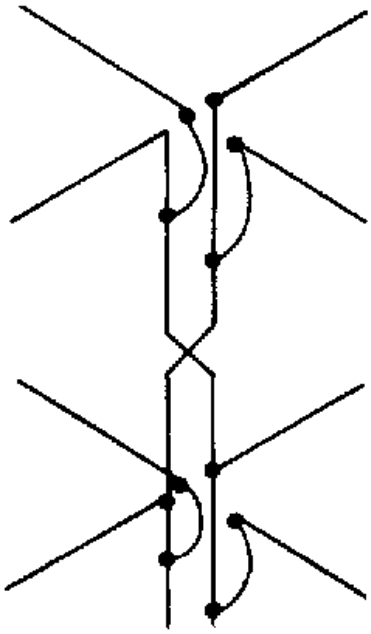
A **V ANTENNA** is a bi-directional antenna consisting of two horizontal, long wires arranged to form a V.



The **RHOMBIC ANTENNA** uses four conductors joined to form a rhombus shape. This antenna has a wide frequency range, is easy to construct and maintain, and is noncritical as far as operation and adjustment are concerned.



The **TURNSTILE ANTENNA** consists of two horizontal, half-wire antennas mounted at right angles to each other.



ANSWERS TO QUESTIONS Q1. THROUGH Q48.

A1. Half-wave (Hertz) and quarter-wave (Marconi).

A2. Coupling device, feeder, and antenna.

A3. Frequency of operation of the transmitter, amount of power to be radiated, and general direction of the receiving set.

A4. One-half the wavelength.

A5. Current and voltage loops.

A6. Current and voltage nodes.

A7. Reciprocity of antennas.

A8. Electric (E) field.

A9. Circular polarization.

A10. Vertical polarization.

A11. Less interference is experienced by man-made noise sources.

A12. Vertical polarization.

A13. 73 ohms.

A14. Anisotropic radiator.

A15. Isotropic radiator.

A16. Anisotropic radiator.

A17. Dipole, doublet and Hertz.

A18. Nondirectional.

A19. Vertical plane.

A20. The pattern would flatten.

A21. To connect one end through a capacitor to the final output stage of the transmitter.

A22. A circular radiation pattern in the horizontal plane, or same as a half wave.

A23. It is composed of a series of conductors arranged in a radial pattern and buried 1 to 2 feet below the ground.

A24. Nine times the feed-point impedance.

A25. Folded dipole.

A26. To produce desired phase relationship between connected elements.

A27. Major lobes have the greatest amount of radiation.

A28. Four.

A29. As more elements are added, an unbalanced condition in the system occurs which impairs efficiency.

A30. By increasing the lengths of the elements of the array.

A31. Directivity increases.

A32. Lower radiation resistance.

A33. Parallel and in the same plane.

A34. They sharpen.

A35. Extremely low radiation resistance, confined to one frequency, and affected by atmospheric conditions.

A36. Along the major axis

A37. Symmetrically.

A38. Length of the parasitic element (tuning) and spacing between the parasitic and driven elements.

A39. Increased gain and directivity.

A40. Rotary array.

A41. Their adjustment is critical and they do not operate over a wide frequency range.

A42. Increased gain.

A43. Multielement parasitic array.

A44. One-half wavelength.

A45. Wave antenna.

A46. Opposite.

A47. It requires a large antenna site.

A48. For omnidirectional vhf communications.

APPENDIX I

GLOSSARY

ABSORPTION—(1) Absorbing light waves. Does not allow any reflection or refraction. (2) Atmospheric absorption of rf energy with no reflection or refraction (adversely affects long distance communications).

ACOUSTICS—The science of sound.

AMPLITUDE—The portion of a cycle measured from a reference line to a maximum value above (or to a maximum value below) the line.

ANGLE OF INCIDENCE—The angle between the incident wave and the normal.

ANGLE OF REFLECTION—The angle between the reflected wave and the normal.

ANGLE OF REFRACTION—The angle between the normal and the path of a wave through the second medium.

ANGSTROM UNIT—The unit used to define the wavelength of light waves.

ANISOTROPIC—The property of a radiator to emit strong radiation in one direction.

ANTENNA—A conductor or set of conductors used either to radiate rf energy into space or to collect rf energy from space.

ARRAY OF ARRAYS—Same as COMBINATION ARRAY.

BAY—Part of an antenna array.

BEVERAGE ANTENNA—A horizontal, longwire antenna designed for reception and transmission of low-frequency, vertically polarized ground waves.

BIDIRECTIONAL ARRAY—An array that radiates in opposite directions along the line of maximum radiation.

BROADSIDE ARRAY—An array in which the direction of maximum radiation is perpendicular to the plane containing the elements.

CENTER-FEED METHOD—Connecting the center of an antenna to a transmission line, which is then connected to the final (output) stage of the transmitter.

CHARACTERISTIC IMPEDANCE—The ratio of voltage to current at any given point on a transmission line. Represented by a value of impedance.

COAXIAL LINE—A type of transmission line that contains two concentric conductors.

COLLINEAR ARRAY—An array with all the elements in a straight line. Maximum radiation is perpendicular to the axis of the elements.

COMBINATION ARRAY—An array system that uses the characteristics of more than one array.

COMPLEMENTARY (SECONDARY) COLORS OF LIGHT—The colors of light produced when two of the primaries are mixed in overlapping beams of light. The complementary colors of light are magenta, yellow, and cyan.

COMPLEX WAVE—A wave produced by combining two or more pure tones at the same time.

COMPRESSION WAVES—Longitudinal waves that have been compressed (made more dense) as they move away from the source.

CONDUCTANCE—The opposite of resistance in transmission lines. The minute amount of resistance that is present in the insulator of a transmission line.

CONNECTED ARRAY—Another term for DRIVEN ARRAY.

COPPER LOSSES—The I^2R loss in a conductor caused by the current flow through the resistance of the conductor.

CORNER-REFLECTOR ANTENNA—A half-wave antenna with a reflector consisting of two flat metal surfaces meeting at an angle behind the radiator.

COUNTERPOISE—A network of wire that is connected to a quarter-wave antenna at one end and provides the equivalent of an additional 1/4 wavelength.

COUPLING DEVICE—A coupling coil that connects the transmitter to the feeder.

CREST (TOP)—The peak of the positive alternation (maximum value above the line) of a wave.

CRITICAL ANGLE—The maximum angle at which radio waves can be transmitted and still be refracted back to earth.

CRITICAL FREQUENCY—The maximum frequency at which a radio wave can be transmitted vertically and still be refracted back to earth.

CURRENT-FEED METHOD—Same as CENTER-FEED METHOD.

CURRENT STANDING-WAVE RATIO (ISWR)—The ratio of maximum to minimum current along a transmission line.

CYCLE—One complete alternation of a sine wave that has a maximum value above and a maximum value below the reference line.

DAMPING—Reduction of energy by absorption.

DENSITY—(1) The compactness of a substance. (2) Mass per unit volume.

DETECTOR—The device that responds to a wave or disturbance.

DIELECTRIC HEATING—The heating of an insulating material by placing it in a high frequency electric field.

DIELECTRIC LOSSES—The losses resulting from the heating effect on the dielectric material between conductors.

DIFFRACTION—The bending of the paths of waves when the waves meet some form of obstruction.

DIFFUSION—The scattering of reflected light waves (beams) from an object, such as white paper.

DIPOLE—A common type of half-wave antenna made from a straight piece of wire cut in half. Each half operates at a quarter wavelength of the output.

DIRECTIONAL—Radiation that varies with direction.

DIRECTOR—The parasitic element of an array that reinforces energy coming from the driver toward itself.

DIRECTIVITY—The property of an array that causes more radiation to take place in certain directions than in others.

DISPERSION—The refraction of light waves that causes the different frequencies to bend at slightly different angles.

DISTRIBUTED CONSTANTS—The constants of inductance, capacitance, and resistance in a transmission line. The constants are spread along the entire length of the line and cannot be distinguished separately.

DOPPLER EFFECT—The apparent change in frequency or pitch when a sound source moves either toward or away from a listener.

DOUBLET—Another name for the dipole antenna.

DRIVEN ARRAY—An array in which all of the elements are driven.

DRIVEN ELEMENT—An element of an antenna (transmitting or receiving) that is connected directly to the transmission line.

ECHO—The reflection of the original sound wave as it bounces off a distant surface.

ELASTICITY—The ability of a substance to return to its original state.

ELECTROMAGNETIC FIELD—The combination of an electric (E) field and a magnetic (H) field.

ELECTROMAGNETIC INTERFERENCE—Man-made or natural interference that degrades the quality of reception of radio waves.

ELECTROMAGNETIC RADIATION—The radiation of radio waves into space.

ELECTRIC (E) FIELD—The field produced as a result of a voltage charge on a conductor or antenna.

ELEMENT—A part of an antenna that can be either an active radiator or a parasitic radiator.

END-FEED METHOD—Connecting one end of an antenna through a capacitor to the final output stage of a transmitter.

END-FIRE ARRAY—An array in which the direction of radiation is parallel to the axis of the array.

FADING—Variations in signal strength by atmospheric conditions.

FEEDER—A transmission line that carries energy to the antenna.

FLAT LINE—A transmission line that has no standing waves. This line requires no special tuning device to transfer maximum power.

FLEXIBLE COAXIAL LINE—A coaxial line made with a flexible inner conductor insulated from the outer conductor by a solid, continuous insulating material.

FOLDED DIPOLE—An ordinary half-wave antenna (dipole) that has one or more additional conductors connected across the ends parallel to each other.

FOUR-ELEMENT ARRAY—An array with three parasitic elements and one driven element.

FREE-SPACE LOSS—The loss of energy of a radio wave because of the spreading of the wavefront as it travels from the transmitter.

FREQUENCY—The number of cycles that occur in one second. Usually expressed in hertz.

FREQUENCY DIVERSITY—Transmitting (and receiving) of radio waves on two different frequencies simultaneously.

FRONT-TO-BACK RATIO—The ratio of the energy radiated in the principal direction to the energy radiated in the opposite direction.

FUNDAMENTAL FREQUENCY—The basic frequency or first harmonic frequency.

GAIN—The ratio between the amount of energy propagated from an antenna that is directional to the energy from the same antenna that would be propagated if the antenna were not directional.

GENERATOR END—See INPUT END.

GROUND PLANE—The portion of a groundplane antenna that acts as ground.

GROUND-PLANE ANTENNA—A type of antenna that uses a ground plane as a simulated ground to produce low-angle radiation.

GROUND REFLECTION LOSS—The loss of rf energy each time a radio wave is reflected from the Earth's surface.

GROUND SCREEN—A series of conductors buried below the surface of the earth and arranged in a radial pattern. Used to reduce losses in the ground.

GROUND WAVES—Radio waves that travel near the surface of the Earth.

HALF-WAVE DIPOLE ANTENNA—An antenna consisting of two rods (1/4 wavelength each) in a straight line, that radiates electromagnetic energy.

HARMONIC—A frequency that is a whole number multiple of a smaller base frequency.

HERTZ ANTENNA—A half-wave antenna installed some distance above ground and positioned either vertically or horizontally.

HORIZONTAL AXIS—On a graph, the straight line axis plotted from left to right.

HORIZONTAL PATTERN—The part of a radiation pattern that is radiated in all directions along the horizontal plane.

HORIZONTALLY POLARIZED—Waves that are radiated with their E field component parallel to the Earth's surface.

INCIDENT WAVE—(1) The wave that strikes the surface of a medium. (2) The wave that travels from the sending end to the receiving end of a transmission line.

INDUCTION FIELD—The electromagnetic field produced about an antenna when current and voltage are present on the same antenna.

INDUCTION LOSSES—The losses that occur when the electromagnetic field around a conductor cuts through a nearby metallic object and induces a current into that object.

INFRASONIC (SUBSONIC)—Sounds below 15 hertz.

INPUT END—The end of a two-wire transmission line that is connected to a source.

INPUT IMPEDANCE—The impedance presented to the transmitter by the transmission line and its load.

INTENSITY (OF SOUND)—The measurement of the amplitude of sound energy. Sometimes mistakenly called loudness.

INTERCEPT—The point where two lines drawn on a graph cross each other.

INTERFERENCE—Any disturbance that produces an undesirable response or degrades a wave.

IONOSPHERE—The most important region of the atmosphere extending from 31 miles to 250 miles above the earth. Contains four cloud-like layers that affect radio waves.

IONOSPHERIC STORMS—Disturbances in the earth's magnetic field that make communications practical only at lower frequencies.

IONIZATION—The process of upsetting electrical neutrality.

ISOTROPIC RADIATION—The radiation of energy equally in all directions.

LEAKAGE CURRENT—The small amount of current that flows between the conductors of a transmission line through the dielectric.

LIGHT RAYS—Straight lines that represent light waves emitting from a source.

LOAD END—See OUTPUT END.

LOADING—See LUMPED-IMPEDANCE TUNING.

LOBE—An area of a radiation pattern plotted on a polar-coordinate graph that represents maximum radiation.

LONG-WIRE ANTENNA—An antenna that is a wavelength or more long at its operating frequency.

LONGITUDINAL WAVES—Waves in which the disturbance (back and forth motion) takes place in the direction of propagation. Sometimes called compression waves.

LOOP—The curves of a standing wave or antenna that represent amplitude of current or voltage.

LOWEST USABLE FREQUENCY—The minimum operating frequency that can be used for communications between two points.

LUMPED CONSTANTS—The properties of inductance, capacitance, and resistance in a transmission line.

LUMPED-IMPEDANCE TUNING—The insertion of an inductor or capacitor in series with an antenna to lengthen or shorten the antenna electrically.

MAGNETIC (H) FIELD—The field produced when current flows through a conductor or antenna.

MAJOR LOBE—The lobe in which the greatest amount of radiation occurs.

MARCONI ANTENNA—A quarter-wave antenna oriented perpendicular to the earth and operated with one end grounded.

MAXIMUM USABLE FREQUENCY— Maximum frequency that can be used for communications between two locations for a given time of day and a given angle of incidence.

MEDIUM—The substance through which a wave travels from one point to the next. Air, water, wood, etc., are examples of a medium.

MINOR LOBE—The lobe in which the radiation intensity is less than a major lobe.

MULTIELEMENT ARRAY—An array consisting of one or more arrays and classified as to directivity.

MULTIELEMENT PARASITIC ARRAY— An array that contains two or more parasitic elements and a driven element.

MULTIPATH—The multiple paths a radio wave may follow between transmitter and receiver.

NATURAL HORIZON—The line-of-sight horizon.

NEGATIVE ALTERNATION—The portion of a sine wave below the reference line.

NODE—The fixed minimum points of voltage or current on a standing wave or antenna.

NOISE (OF SOUND)—An unwanted disturbance caused by spurious waves that originate from man-made or natural sources.

NONDIRECTIONAL—See OMNIDIRECTIONAL.

NONLUMINOUS BODIES—Objects that either reflect or diffuse light that falls upon them.

NONRESONANT LINE—A transmission line that has no standing waves of current or voltage.

NORMAL—The imaginary line perpendicular to the point at which the incident wave strikes the reflecting surface. Also called the perpendicular.

NULL—On a polar-coordinate graph, the area that represents minimum or 0 radiation.

OMNIDIRECTIONAL—Transmitting in all directions.

OPAQUE—A type of substance that does not transmit any light rays.

OPEN-ENDED LINE—A transmission line that has an infinitely large terminating impedance.

OPTIMUM WORKING FREQUENCY—The most practical operating frequency that can be used with the least amount of problems; roughly 85 percent of the maximum usable frequency.

ORIGIN—The point on a graph where the vertical and horizontal axes cross each other.

OUTPUT END—The end of a transmission line that is opposite the source.

OUTPUT IMPEDANCE—The impedance presented to the load by the transmission line and its source.

PARALLEL RESONANT CIRCUIT—A circuit that acts as a high impedance at resonance.

PARALLEL-WIRE—A type of transmission line consisting of two parallel wires.

PARASITIC ARRAY—An array that has one or more parasitic elements.

PARASITIC ELEMENT—The passive element of an antenna array that is connected to neither the transmission line nor the driven element.

PERIOD—The amount of time required for completion of one full cycle.

PITCH—A term used to describe the frequency of a sound heard by the human ear.

PLANE OF POLARIZATION—The plane (vertical or horizontal) with respect to the earth in which the E field propagates.

POINT OF ZERO DISPLACEMENT—See REFERENCE LINE.

POLAR-COORDINATE GRAPH—A graph whose axes consist of a series of circles with a common center and a rotating radius extending from the center of the concentric circles.

POSITIVE ALTERNATION—The portion of a sine wave above the reference line.

POWER LOSS—The heat loss in a conductor as current flows through it.

POWER STANDING-WAVE RATIO (PSWR)—The ratio of the square of the maximum and minimum voltages of a transmission line.

PRIMARY COLORS (OF LIGHT)—The three primary colors of light (red, green, and blue), from which all other colors may be derived.

PRISM—A triangular-shaped glass that refracts and disperses light waves into component wavelengths.

PROPAGATION—Waves traveling through a medium.

QUALITY (OF SOUND)—The factor that distinguishes tones of pitch and loudness.

QUARTER-WAVE ANTENNA—Same as the Marconi antenna.

RADIATION FIELD—The electromagnetic field that detaches itself from an antenna and travels through space.

RADIATION LOSSES—The losses that occur when magnetic lines of force about a conductor are projected into space as radiation and are not returned to the conductor as the cycle alternates.

RADIATION PATTERN—A plot of the radiated energy from an antenna.

RADIATION RESISTANCE—The resistance, which if inserted in place of an antenna, would consume the same amount of power as that radiated by the antenna.

RADIO FREQUENCIES—Electromagnetic frequencies that fall between 3 kilohertz and 300 gigahertz and are used for radio communications.

RADIO HORIZON—The boundary beyond the natural horizon in which radio waves cannot be propagated over the earth's surface.

RADIO WAVE—(1) A form of radiant energy that can neither be seen nor felt. (2) An electromagnetic wave generated by a transmitter.

RAREFIED WAVE—A longitudinal wave that has been expanded or rarefied (made less dense) as it moves away from the source.

RECEIVER—The object that responds to a wave or disturbance. Same as detector.

RECEIVING ANTENNA—The device used to pick up an rf signal from space.

RECEIVING END—See OUTPUT END.

RECIPROCITY—The property of interchangeability of the same antenna for transmitting and receiving.

RECTANGULAR-COORDINATE GRAPH—A graph in which straight-line axes (horizontal and vertical) are perpendicular.

REFERENCE LINE—The position a particle of matter would occupy if it were not disturbed by wave motion.

REFLECTED WAVE—(1) The wave that reflects back from a medium. (2) Waves traveling from the load back to the generator on a transmission line. (3) The wave moving back to the sending end of a transmission line after reflection has occurred.

REFLECTION WAVES—Waves that are neither transmitted nor absorbed, but are reflected from the surface of the medium they encounter.

REFLECTOR—The parasitic element of an array that causes maximum energy radiation in a direction toward the driven element.

REFRACTION—The changing of direction as a wave leaves one medium and enters another medium of a different density.

RERADIATION—The reception and retransmission of radio waves caused by turbulence in the troposphere.

RESONANCE—The condition produced when the frequency of vibrations are the same as the natural frequency (of a cavity). The vibrations reinforce each other.

RESONANT LINE—A transmission line that has standing waves of current and voltage.

REST POSITION—See REFERENCE LINE.

REVERBERATION—The multiple reflections of sound waves.

RHOMBIC ANTENNA—A diamond-shaped antenna used widely for long-distance, high-frequency transmission and reception.

RIGID COAXIAL LINE—A coxial line consisting of a central, insulated wire (inner conductor) mounted inside a tubular outer conductor.

SCATTER ANGLE—The angle at which the receiving antenna must be aimed to capture the scattered energy of tropospheric scatter.

SELF-INDUCTION—The phenomenon caused by the expanding and collapsing fields of an electron which encircles other electrons and retards the movement of the encircled electrons.

SELF-LUMINOUS BODIES—Objects that produce their own light.

SENDING END—See INPUT END.

SERIES RESONANT CIRCUIT—A circuit that acts as a low impedance at resonance.

SHIELDED PAIR—A line consisting of parallel conductors separated from each other and surrounded by a solid dielectric.

SHORT-CIRCUITED LINE—A transmission line that has a terminating impedance equal to 0.

SINK—See OUTPUT END.

SKIN EFFECT—The flow of ac current near the surface of a conductor at rf frequencies.

SKIP DISTANCE—The distance from a transmitter to the point where the sky wave is first returned to earth.

SKIP ZONE—A zone of silence between the point where the ground wave becomes too weak for reception and the point where the sky wave is first returned to earth.

SKY WAVES—Radio waves reflected back to earth from the ionosphere.

SONIC—Pertaining to sounds capable of being heard by the human ear.

SOURCE—(1) The object that produces waves or disturbance. (2) The name given to the end of a two-wire transmission line that is connected to a source.

SPACE DIVERSITY—Reception of radio waves by two or more antennas spaced some distance apart.

SPACE WAVE—A radio wave that travels directly from the transmitter to the receiver and remains in the troposphere.

SPECTRUM—(1) The entire range of electromagnetic waves. (2) VISIBLE. The range of electromagnetic waves that stimulate the sense of sight. (3) ELECTROMAGNETIC. The entire range of electromagnetic waves arranged in order of their frequencies.

SPORADIC E LAYER—Irregular cloud-like patches of unusually high ionization. Often forms at heights near the normal E layer.

SPREADER—Insulator used with transmission lines and antennas to keep the parallel wires separated.

STANDING WAVE—The distribution of voltage and current formed by the incident and reflected waves which have minimum and maximum points on a resultant wave that appears to stand still.

STANDING-WAVE RATIO (SWR)—The ratio of the maximum (voltage, current) to the minimum (voltage, current) of a transmission line. Measures the perfection of the termination of the line.

STRATOSPHERE—Located between the troposphere and the ionosphere. Has little effect on radio waves.

STUB—Short section of a transmission line used to match the impedance of a transmission line to an antenna. Can also be used to produce desired phase relationships between connected elements of an antenna.

SUDDEN IONOSPHERIC DISTURBANCE—An irregular ionospheric disturbance that can totally blank out hf radio communications.

SUPERSONIC—Speed greater than the speed of sound.

SURFACE WAVE—A radio wave that travels along the contours of the earth, thereby being highly attenuated.

TEMPERATURE INVERSION—The condition in which warm air is formed above a layer of cool air that is near the earth's surface.

THREE-ELEMENT ARRAY—An array with two parasitic elements (reflector and director) and a driven element.

TONES—Musical sounds.

TRANSLUCENT—A type of substance, such as frosted glass, through which some light rays can pass but through which objects cannot be seen clearly.

TRANSMISSION LINE—A device designed to guide electrical energy from one point to another.

TRANSMITTING ANTENNA—The device used to send the transmitted signal energy into space.

TRANSPARENT—A type of substance, such as glass, that transmits almost all of the light waves that fall upon it.

TRANSMISSION MEDIUMS—The various types of lines and waveguides used as transmission lines.

TRANSMITTER END—See INPUT END.

TRANSVERSE WAVE MOTION—The up and down motion of a wave as the wave moves outward.

TROPOSPHERE—The portion of the atmosphere closest to the earth's surface, where all weather phenomena take place.

TROPOSPHERIC SCATTER—The propagation of radio waves in the troposphere by means of scatter.

TROUGH (BOTTOM)—The peak of the negative alternation (maximum value below the line).

TUNED LINE—Another name for the resonant line. This line uses tuning devices to eliminate the reactance and to transfer maximum power from the source to the line.

TURNSTILE ANTENNA—A type of antenna used in vhf communications that is omnidirectional and consists of two horizontal half-wave antennas mounted at right angles to each other in the same horizontal plane.

TWISTED PAIR—A line consisting of two insulated wires twisted together to form a flexible line without the use of spacers.

TWO-WIRE OPEN LINE—A parallel line consisting of two wires that are generally spaced from 2 to 6 inches apart by insulating spacers.

TWO-WIRE RIBBON (TWIN LEAD)—A parallel line similar to a two-wire open line except that uniform spacing is assured by embedding the two wires in a low-loss dielectric.

ULTRASONIC—Sounds above 20,000 hertz.

UNIDIRECTIONAL ARRAY—An array that radiates in only one general direction.

UNTUNED LINE—Another name for the flat or nonresonant line.

V ANTENNA—A bi-directional antenna, shaped like a V, which is widely used for communications.

VELOCITY—The rate at which a disturbance travels through a medium.

VERTICAL AXIS—On a graph, the straight line axis oriented from bottom to top.

VERTICAL PATTERN—The part of a radiation pattern that is radiated in the vertical plane.

VERTICALLY POLARIZED—Waves radiated with the E field component perpendicular to the earth's surface.

VOLTAGE-FEED METHOD—Same as END FEED METHOD.

VOLTAGE STANDING-WAVE RATIO (VSWR)—The ratio of maximum to minimum voltage of a transmission line.

WAVE ANTENNA—Same as BEVERAGE ANTENNA.

WAVE MOTION—A recurring disturbance advancing through space with or without the use of a physical medium.

WAVE TRAIN—A continuous series of waves with the same amplitude and wavelength.

WAVEFRONT—A small section of an expanding sphere of electromagnetic radiation, perpendicular to the direction of travel of the energy.

WAVEGUIDE—A hollow metal tube used as a transmission line to guide energy from one point to another.

WAVELENGTH—(1) The distance in space occupied by 1 cycle of a radio wave at any given instant. (2) The distance a disturbance travels during one period of vibration.

YAGI ANTENNA—A multielement parasitic array. Elements lie in the same plane as those of the end-fire array.

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Assignment Questions

Information: The text pages that you are to study are provided at the beginning of the assignment questions.

ASSIGNMENT 1

Textbook assignment: Chapter 1, “Wave Propagation,” pages 1-1 through 1-48.

1-1. What is the major advantage of the telegraph over earlier methods of communication?

1. Range
2. Speed
3. Security
4. Reliability

1-2. The spreading out of radio waves is referred to as propagation and is used in which of the following Navy equipment?

1. Detection
2. Communication
3. Radar and navigation
4. Each of the above

1-3. Radio-frequency waves CANNOT be seen for which of the following reasons?

1. Because radio-frequency energy is low powered
2. Because radio-frequency waves are below the sensitivity range of the human eye
3. Because the human eye detects only magnetic energy
4. Because radio-frequency waves are above the sensitivity range of the human eye

1-4. Radio waves travel at what speed?

1. Speed of sound
2. Speed of light
3. Speed of the Earth's rotation
4. Speed of the Earth's orbit around the sun

1-5. Which of the following types of energy CANNOT be seen, heard, or felt?

1. Radio waves
2. Sound waves
3. Heat waves
4. Light waves

1-6. A stone dropped into water creates a series of expanding circles on the surface of the water. This is an example of which of the following types of wave motion?

1. Transverse
2. Concentric
3. Longitudinal
4. Compression

1-7. A sound wave that moves back and forth in the direction of propagation is an example of which of the following types of wave motion?

1. Composite
2. Concentric
3. Transverse
4. Longitudinal

1-8. Which of the following terms is used for the vehicle through which a wave travels from point to point?

1. Medium
2. Source
3. Detector
4. Receiver

1-9. Which of the following is NOT an element necessary to propagate sound?

1. Medium
2. Source
3. Detector
4. Reference

1-10. If a wave has a velocity of 4,800 feet per second and a wave-length of 5 feet, what is the frequency of the wave?

1. 9.6 Hz
2. 96 Hz
3. 960 Hz
4. 9,600 Hz

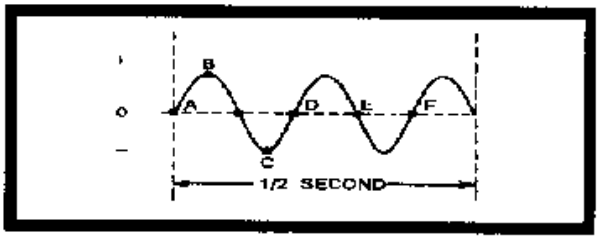

Figure 1-A.—Waveform.

IN ANSWERING QUESTIONS 1-11 THROUGH 1-15, REFER TO FIGURE 1-A.

1-11. The waveform in the figure is what type of wave?

1. Sine
2. Square
3. Sawtooth
4. Trapezoidal

1-12. The distance between which of the following points represents the completion of a full cycle of alternating current?

1. A to C
2. B to D
3. C to E
4. D to F

1-13. The distance between which of the following points represents a full wavelength?

1. A to D
2. A to E
3. D to E
4. E to F

1-14. What is the frequency of the wave?

1. 0.5 Hz
2. 2.5 Hz
3. 5.0 Hz
4. 7.5 Hz

1-15. What is the period of the wave?

1. 100 milliseconds
2. 200 milliseconds
3. 250 milliseconds
4. 500 milliseconds

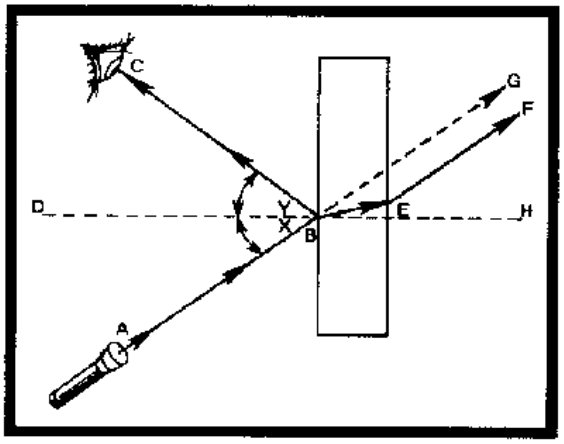

Figure 1-B.—Wave angles.

IN ANSWERING QUESTIONS 1-16 THROUGH 1-19, REFER TO FIGURE 1-B.

1-16. What line in the figure indicates the incident wave?

1. A to B
2. B to E
3. C to B
4. D to H

1-17. Angle "x" is which of the following angles?

1. Normal
2. Incidence
3. Reflection
4. Refraction

1-18. Line E to F represents which of the following waves?

1. Normal
2. Incident
3. Refracted
4. Reflected

1-19. Line D to H represents which of the following references?

1. Normal
2. Perpendicular
3. Both 1 and 2 above
4. Reflected line

1-20. Which of the following statements about a wave is the law of reflection?

1. The angle of incidence is equal to the refracted wave
2. The angle of incidence is not equal to the refracted wave
3. The angle of incidence is equal to the angle of reflection
4. The angle of incidence is not equal to the angle of reflection

1-21. If a wave passes first through a dense medium and then through a less dense medium, which of the following angle-of-refraction conditions exists?

1. The angle of refraction is greater than the angle of incidence
2. The angle of refraction is less than the angle of incidence
3. The angle of refraction is equal to the angle of incidence
4. The wave will pass through in a straight line

1-22. The reception of an AM-band radio signal over mountains can be explained by which of the following principles of wave propagation?

1. Reflection
2. Refraction
3. Diffraction
4. Doppler effect

1-23. What wave propagation principle accounts for the apparent increase in frequency as a train whistle approaches and the apparent decrease in frequency as it moves away?

1. Refraction
2. Reflection
3. Diffraction
4. Doppler effect

1-24. Longitudinal wave disturbances that travel through a medium are known as what type of waves?

1. Air
2. Sound
3. Radio
4. Light

1-25. What are the three audible frequency ranges?

1. Subsonic, sonic, and supersonic
2. Infrasonic, sonic, and ultrasonic
3. Infrasonic, subsonic, and ultrasonic
4. Infrasonic, subsonic, and supersonic

1-26. If a bell is placed in a jar and the air in the jar is replaced with a gas of a higher density, what is the effect, if any, on the speed of the sound when the bell is rung?

1. The sound stops
2. The sound travels faster
3. The sound travels slower
4. The sound is not affected

1-27. Varying which of the following wave characteristics will cause the length of sound waves to vary?

1. Phase
2. Quality
3. Amplitude
4. Frequency

1-28. What are the three basic characteristics of sound?

1. Amplitude, intensity, and quality
2. Amplitude, pitch, and tone
3. Pitch, intensity, and quality
4. Pitch, frequency, and quality

1-29. If several musical instruments are playing the same note, you should be able to distinguish one instrument from another because of which of the following characteristics of sound?

1. Quality
2. Overtones
3. Frequency
4. Intensity

1-30. Through which of the following mediums will sound travel fastest, at the indicated temperature?

1. Air at 68º F
2. Lead at 20º C
3. Steel at 32º F
4. Steel at 20º C

1-31. In sound terminology, which of the following terms is the same as a wave reflection?

1. Echo
2. Image
3. Acoustics
4. Refraction

1-32. Multiple reflections of sound waves are referred to as

1. noise
2. acoustics
3. interference
4. reverberation

1-33. Two out-of-phase waves of the same frequency that are moving through the same medium are said to present which of the following types of interference?

1. Additive
2. Constructive
3. Both 1 and 2 above
4. Subtractive

1-34. A cavity that vibrates at its own natural frequency and produces a sound that is louder than at other frequencies is demonstrating which of the following sound characteristics?

1. Noise
2. Quality
3. Resonance
4. Reverberation

1-35. Energy in the form of light can be produced through which of the following means?

1. Chemical
2. Electrical
3. Mechanical
4. Each of the above

1-36. The scientist, J. C. Maxwell, developed the theory that small packets of electromagnetic energy called photons produce

1. sound
2. noise
3. echoes
4. light

1-37. A large volume of light radiating in a given direction is referred to as a

1. ray
2. beam
3. shaft
4. pencil

1-38. Which of the following units of measurement is/are used to measure very short wavelengths of light?

1. Angstrom (Å)
2. Millimicron
3. Both 1 and 2 above
4. Millimeter

1-39. What are the primary colors of light?

1. Red, blue, and yellow
2. Red, blue, and green
3. Red, violet, and indigo
4. Blue, green, and violet

1-40. What are the secondary colors of light?

1. Orange, yellow, and blue-green
2. Magenta, yellow, and cyan
3. Purple, yellow, and black
4. Red, white, and blue

1-41. What causes sunlight to separate into different wavelengths and display a rainbow of colors when passed through a prism?

1. Refraction
2. Reflection
3. Dispersion
4. Diffraction

1-42. The sun, gas flames, and electric light filaments are visible because they are

1. opaque
2. transparent
3. nonluminous
4. self-luminous

1-43. Substances that transmit almost all of the light waves falling upon them possess which of the following properties?

1. Opaqueness
2. Transparence
3. Translucence
4. Self-lumination

1-44. Some substances are able to transmit light waves but objects cannot be seen through them. Which of the following properties does this statement describe?

1. Opaqueness
2. Transparence
3. Translucence
4. Self-lumination

1-45. The speed of light depends on the medium through which light travels. For which of the following reasons does light travel through empty space faster than through an object such as glass?

1. Space is less dense than glass
2. Space is more dense than glass
3. Glass reflects the light back to the source
4. Glass refracts the light, causing the light to travel in all directions

1-46. If a light wave strikes a sheet of glass at a perpendicular angle, what is the effect, if any, on the light wave?

1. The wave is completely absorbed
2. The wave is reflected back toward the source
3. The wave is refracted as it passes through the glass
4. The wave is unchanged and continues in a straight line

1-47. The amount of absorption of the light that strikes an object is determined by the object's

1. color
2. purity
3. density
4. complexity

1-48. In a comparison of waves of light and sound as they travel from an air into water, how is the speed of (a) light waves and (b) sound waves affected?

1. (a) Increased (b) increased
2. (a) Increased (b) decreased
3. (a) Decreased (b) decreased
4. (a) Decreased (b) increased

1-49. Which of the following waves are NOT a form of electromagnetic energy?

1. Heat waves
2. Sound waves
3. Light waves
4. Radio waves

1-50. The electromagnetic spectrum represents the entire range of electromagnetic waves arranged in the order of their

1. color
2. frequency
3. visibility
4. application

1-51. Which of the following portions of the frequency spectrum contains the highest frequency?

1. X-ray
2. Radar
3. Light
4. Cosmic

1-52. Which of the following electronic devices is used to radiate and/or collect electromagnetic waves?

1. Antenna
2. Receiver
3. Transmitter
4. Transmission line

1-53. The electric field and magnetic field combine to form which of the following types of waves?

1. Spherical
2. Elliptical
3. Electromagnetic
4. Each of the above

1-54. The magnetic field radiated from an antenna is produced by what electrical property?

1. Voltage
2. Current
3. Reactance
4. Resistance

1-55. The electric field radiated from an antenna is produced by what electrical property?

1. Voltage
2. Current
3. Reactance
4. Resistance

1-56. Applying rf energy to the elements of an antenna results in what phase relationship between voltage and current?

1. Voltage lags current by 90 degrees
2. Voltage leads current by 90 degrees
3. Voltage and current are 180 degrees out of phase
4. Voltage and current are in phase

1-57. What field exists close to the conductor of an antenna and carries the current?

1. Electric
2. Magnetic
3. Induction
4. Radiation

1-58. What field travels through space after being detached from the current-carrying rod of an antenna?

1. Electric
2. Magnetic
3. Induction
4. Radiation

1-59. Electric and magnetic fields on an antenna reach their maximum intensity at which of the following times?

1. When they are a full cycle apart
2. When they are three-quarter cycle apart
3. When they are a half-cycle apart
4. When they are a quarter-cycle apart

ASSIGNMENT 2

Textbook assignment: Chapter 2, "Radio Wave Propagation," pages 2-1 through 2-47.

2-1. The induction field is made up of which of the following fields?

1. E field only
2. H field only
3. Both E and H fields

2-2. After the radiation field leaves an antenna, what is the relationship between the E and H fields with respect to (a) phase and (b) physical displacement in space?

1. (a) In phase (b) 90 degrees
2. (a) Out of phase (b) 90 degrees
3. (a) In phase (b) 180 degrees
4. (a) Out of phase (b) 180 degrees

2-3. What is the first harmonic of a radio wave that has a fundamental frequency of 2,000 kHz?

1. 6,000 kHz
2. 2,000 kHz
3. 3,000 kHz
4. 4,000 kHz

2-4. In a radio wave with a fundamental frequency of 1.5 kHz, which of the following frequencies is NOT a harmonic?

1. 6,000 kHz
2. 5,000 kHz
3. 3,000 kHz
4. 4,000 kHz

2-5. A radio wave with a frequency of 32 kHz is part of which of the following frequency bands?

1. The lf band
2. The mf band
3. The hf band
4. The vhf band

2-6. A frequency of 3.5 GHz falls into what rf band?

1. High
2. Very high
3. Super high
4. Extremely high

2-7. A radio wavelength expressed as 250 meters may also be expressed as how many feet?

1. 410
2. 820
3. 1,230
4. 1,640

2-8. An increase in the frequency of a radio wave will have what effect, if any, on the velocity of the radio wave?

1. Increase
2. Decrease
3. None

2-9. An increase in frequency of a radio wave will have what effect, if any, on the wavelength of the radio wave?

1. Increase
2. Decrease
3. None

2-10. What is the frequency, in kiloHertz, of a radio wave that is 40 meters long?

1. 75
2. 750
3. 7,500
4. 75,000

2-11. What is the approximate wavelength, in feet, of a radio wave with a frequency of 5,000 kHz?

1. 197 feet
2. 1,970 feet
3. 19,700 feet
4. 197,000 feet

2-12. The polarity of a radio wave is determined by the orientation of (a) what moving field with respect to (b) what reference?

1. (a) Electric (b) earth
2. (a) Electric (b) antenna
3. (a) Magnetic (b) antenna
4. (a) Magneti (b) earth

2-13. Energy radiated from an antenna is considered horizontally polarized under which of the following conditions?

1. If the wavefront is in the horizontal plane
2. If the magnetic field is in the horizontal plane
3. If the electric field is in the horizontal plane
4. If the induction field is in the horizontal plane

2-14. The ability of a reflecting surface to reflect a specific radio wave depends on which of the following factors?

1. Striking angle
2. Wavelength of the wave
3. Size of the reflecting area
4. All of the above

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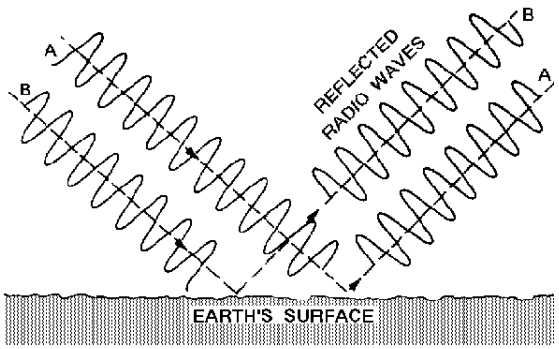

Figure 2-A.—Reflected radio waves.

IN ANSWERING QUESTION 2-15, REFER TO FIGURE 2-A.

2-15. If the two reflected radio waves shown in the figure are received at the same instant at the receiving site, what will be the effect, if any, on signal quality?

1. A stronger signal will be produced
2. A weak or fading signal will be produced
3. The signal will be completely canceled out
4. None

2-16. The bending of a radio wave because of a change in its velocity through a medium is known as

1. refraction
2. reflection
3. deflection
4. diffraction

2-17. Radio communications can be diffracted to exceptionally long distances through the use of (a) what frequency band at (b) what relative power level?

1. (a) Very low frequency (b) Low power
2. (a) Very high frequency (b) Low power
3. (a) Very low frequency (b) High power
4. (a) Very high frequency (b) High power

2-18. Electrically charged particles that affect the propagation of radio waves are found in what atmospheric layer?

1. Troposphere
2. Ionosphere
3. Chronosphere
4. Stratosphere

2-19. Most weather phenomena take place in which of the following region of the atmosphere?

1. Troposphere
2. Ionosphere
3. Chronosphere
4. Stratosphere

2-20. Radio wave propagation has the least effect because of its constancy on which of the following atmospheric layers?

1. Troposphere
2. Ionosphere
3. Chronosphere
4. Stratosphere

2-21. Long range, surface-wave communications are best achieved when the signal is transmitted over seawater with (a) what polarization at (b) what relative frequency?

1. (a) Vertical (b) Low
2. (a) Vertical (b) High
3. (a) Horizontal (b) High
4. (a) Horizontal (b) Low

2-22. The Navy's long-range vlf broadcasts are possible because of the advantages of which of the following types of propagation?

1. Diffraction
2. Ionospheric refraction
3. Repeated reflection and refraction
4. Both 2 and 3 above

2-23. A space wave (a) is primarily a result of refraction in what atmospheric layer and (b) extends approximately what distance beyond the horizon?

1. (a) Ionosphere
   (b) One-tenth farther
2. (a) Ionosphere
   (b) One-third farther
3. (a) Troposphere
   (b) One-third farther
4. (a) Troposphere
   (b) One-tenth farther

2-24. The signal of a space wave is sometimes significantly reduced at the receiving site because of which of the following interactions?

1. Space-wave refraction
2. Space-wave reflections
3. Ground-wave diffraction
4. Ground-wave reflections

2-25. For long-range communications in the hf band, which of the following types of waves is most satisfactory?

1. Sky wave
2. Space wave
3. Surface wave
4. Reflected ground wave

2-26. Ionization in the atmosphere is produced chiefly by which of the following types of radiation?

1. Alpha radiation
2. Cosmic radiation
3. Infrared radiation
4. Ultraviolet radiation

2-27. Ultraviolet waves of higher frequencies produce ionized layers at what relative altitude(s)?

1. Lower
2. Higher
3. Both 1 and 2 above

2-28. The density of ionized layers is normally greatest during which of the following periods?

1. At night
2. Before sunrise
3. Between early morning and late afternoon
4. Between afternoon and sunset

2-29. Compared to the other ionospheric layers at higher altitudes, the ionization density of the D layer is

1. about the same
2. relatively low
3. relatively high

2-30. What two layers in the ionosphere recombine and largely disappear at night?

1. D and F
2. D and E
3. E and F2
4. F1 and F2

2-31. For hf-radio communications covering long distances, what is the most important layer of the ionosphere?

1. C
2. D
3. E
4. F

2-32. Refraction of a sky wave in the ionosphere is influenced by which of the following factors?

1. Ionospheric density
2. Frequency of the wave
3. Angle of incidence of the wave
4. All of the above

2-33. A 10-MHz wave entering the ionosphere at an angle greater than its critical angle will pass through the ionosphere and be lost in space unless which of the following actions is taken?

1. The ground wave is canceled
2. The frequency of the wave is increased
3. The frequency of the wave is decreased
4. The ground wave is reinforced

2-34. The distance between the transmitter and the nearest point at which refracted waves return to earth is referred to as the

1. skip distance
2. return distance
3. reception distance
4. ground-wave distance

2-35. When ground-wave coverage is LESS than the distance between the transmitter and the nearest point at which the refracted waves return to earth, which of the following reception possibilities should you expect?

1. No sky-wave
2. Weak ground wave
3. A zone of silence
4. Strong ground wave

2-36. The greatest amount of absorption takes place in the ionosphere under which of the following conditions?

1. When sky wave intensity is the greatest
2. When collision of particles is least
3. When the density of the ionized layer is the greatest
4. When precipitation is greatest

2-37. Which of the following layers provide the greatest amount of absorption to the ionospheric wave?

1. D and E
2. D and F1
3. E and F1
4. F1 and F2

2-38. If the signal strength of an incoming signal is reduced for a prolonged period, what type of fading is most likely involved?

1. Selective
2. Multipath
3. Absorption
4. Polarization

2-39. Radio waves that arrive at a receiving site along different paths can cause signal fading if these waves have different

1. velocities
2. amplitudes
3. phase relationships
4. modulation percentages

2-40. The technique of reducing multipath fading by using several receiving antennas at different locations is known as what type of diversity?

1. Space
2. Receiver
3. Frequency
4. Modulation

2-41. The amount of rf energy lost because of ground reflections depends on which of the following factors?

1. Angle of incidence
2. Ground irregularities
3. Frequency of the wave
4. Each of the above

2-42. Receiving sites located near industrial areas can expect to have exceptionally large losses in signal quality as a result of which of the following propagation situations?

1. Absorption
2. Multihop refraction
3. Natural interference
4. Man-made interference

2-43. Which of the following ionospheric variation causes densities to vary with the axial rotation of the sun?

1. Daily variation
2. Seasonal variation
3. 27-day sunspot cycle
4. 11-year sunspot cycle

2-44. Which of the following ionospheric variation causes densities to vary with the position of the earth in its orbit around the sun?

1. Daily variation
2. Seasonal variation
3. 27-day sunspot cycle
4. 11-year sunspot cycle

2-45. Which of the following ionospheric variation causes densities to vary with the time of the day?

1. Daily variation
2. Seasonal variation
3. 27-day sunspot cycle
4. 11-year sunspot cycle

2-46. What relative range of operating frequencies is required during periods of maximum sunspot activity?

1. Lower
2. Medium
3. Higher

2-47. What factor significantly affects the frequency of occurrence of the sporadic-E layer?

1. Seasons
2. Latitude
3. Weather conditions
4. Ionospheric storms

2-48. What effect can the sporadic-E layer have on the propagation of sky waves?

1. Causes multipath interference
2. Permits long distance communications at unusually high frequencies
3. Permits short-distance communications in the normal skip zone
4. Each of the above

2-49. A sudden and intense burst of ultraviolet light is especially disruptive to communications in which of the following frequency bands?

1. Hf
2. Mf
3. Lf
4. Vlf

2-50. The density of what ionosphere layer increases because of a violent eruption on the surface of the sun?

1. D
2. E
3. F1
4. F2

2-51. Which irregular variation in ionospheric conditions can cause a waiting period of several days before communications return to normal?

1. Sporadic E
2. Ionospheric storms
3. Sudden ionospheric disturbance
4. Each of the above

2-52. For a radio wave entering the atmosphere of the earth at a given angle, the highest frequency at which refraction will occur is known by which of the following terms?

1. Usable frequency
2. Refraction frequency
3. Maximum usable frequency
4. Optimum working frequency

2-53. The most consistent communications can be expected at which of the following frequencies?

1. Critical frequency
2. Maximum usable frequency
3. Maximum working frequency
4. Optimum working frequency

2-54. If the optimum working frequency for a communications link is 4,250 kHz, what is the approximate maximum usable frequency?

1. 4,500 kHz
2. 5,000 kHz
3. 5,500 kHz
4. 6,000 kHz

2-55. In determining the success of radio transmission, which of the following factors is the LEAST predictable?

1. Antenna capabilities
2. Weather conditions along the path of communication
3. Density of ionized layers
4. Presence of ionized layers

2-56. At frequencies above 100 MHz, the greatest attenuation of rf energy from raindrops is caused by which of the following factors?

1. Ducting
2. Heat loss
3. Scattering
4. Absorption

2-57. Under certain conditions, such as ducting, line-of-sight radio waves often propagate for distances far beyond their normal ranges because of which of the following factors?

1. Low cloud masses
2. Ionospheric storms
3. Temperature inversions
4. Frequency fluctuations

2-58. When ducting is present in the atmosphere, multihop refraction of line-of-sight transmission can occur because of which of the following factors?

1. Operating frequency of the transmitter
2. Height of the transmitting antenna
3. Angle of incidence of the radio wave
4. Each of the above

2-59. A propagation technique used to extend uhf transmission range beyond the horizon uses which of the following propagation characteristics?

1. Ground reflection
2. Ionospheric scatter
3. Tropospheric scatter
4. Atmospheric refraction

2-60. Communications by tropospheric scatter can be affected by which of the following conditions?

1. Sunspot activity
2. Atmospheric conditions
3. Ionospheric disturbances
4. All of the above

2-61. What effect, if any, does the radiation angle of a transmitting antenna have on the reception of communications by tropospheric scatter?

1. The lower the angle, the weaker the signal
2. The lower the angle, the stronger the signal
3. The lower the angle, the more susceptible the signal is to distortion
4. None

2-62. Which of the following descriptions of tropospheric scatter signal reception is NOT true?

1. Receiver signal strength decreases as the turbulence height is increased
2. The level of reception depends on the number of turbulences causing scatter
3. The energy received is the portion of the wave reradiated by the turbulence
4. Increased communications distance enables more turbulence to act on the signal, thereby raising the received signal level

2-63. The tropospheric scatter signal is often characterized by very rapid fading caused by which of the following factors?

1. Extreme path lengths
2. Multipath propagation
3. Turbulence in the atmosphere
4. Angle of the transmitted beam

2-64. For which of the following communications situations would turbulence in the troposphere scatter transmission?

1. 10 MHz, range 200 miles
2. 30 MHz, range 800 miles
3. 50 MHz, range 600 miles
4. 100 MHz, range 400 miles

ASSIGNMENT 3

Textbook assignment: Chapter 3, "Principles of Transmission Lines," pages 3-1 through 3-58.

3-1. A transmission line is designed to perform which of the following functions?

1. Disperse energy in all directions
2. Detune a transmitter to match the load
3. Guide electrical energy from point to point
4. Replace the antenna in a communications system

3-2. All transmission lines must have two ends, the input end and the output end. What other name is given to the input end?

1. Sending end
2. Generator end
3. Transmitter end
4. Each of the above

3-3. A measurement of the voltage to current ratio (E_{in}/I_{in}) at the input end of a transmission line is called the

1. input-gain rate
2. input impedance
3. output impedance
4. voltage-gain ratio

3-4. Which of the following lines is NOT a transmission medium?

1. Load line
2. Coaxial line
3. Two-wire line
4. Twisted-pair line

3-5. Electrical power lines are most often made of which of the following types of transmission lines?

1. Twin-lead line
2. Shielded-pair line
3. Two-wire open line
4. Two-wire ribbon line

3-6. Uniform capacitance throughout the length of the line is an advantage of which of the following transmission lines?

1. Coaxial line
2. Twisted pair
3. Shielded pair
4. Two-wire open line

3-7. What is the primary advantage of a rigid coaxial line?

1. Low radiation losses
2. Inexpensive construction
3. Low high-frequency losses
4. Each of the above

3-8. Which of the following wave-guides is seldom used because of its large energy loss characteristics?

1. Metallic
2. Dielectric
3. Elliptical
4. Cylindrical

3-9. To some degree, transmission lines always exhibit which of the following types of losses?

1. I^2R
2. Inductor
3. Dielectric
4. Each of the above

3-10. Skin effect is classified as which of the following types of loss?

1. Copper
2. Voltage
3. Induction
4. Dielectric

3-11. What transmission-line loss is caused by magnetic lines of force not returning to the conductor?

1. Copper
2. Radiation
3. Induction
4. Dielectric

3-12. What is the electrical wave-length of 1 cycle if the frequency is 60 hertz?

1. 125,000 meters
2. 1,250,000 meters
3. 5,000,000 meters
4. 20,000,000 meters

3-13. A transmission line 10 meters in length is considered to be electrically long at which of the following frequencies?

1. 60 kilohertz
2. 600 kilohertz
3. 6 megahertz
4. 60 megahertz

3-14. The conductance value of a transmission line represents which of the following values?

1. Expected value of current flow through the insulation
2. Expected value of voltage supplied by the transmitter
3. Value of the lump and distributed constants of the line divided by impedance
4. Value of the lump and distributed constants of the line divided by impedance

3-15. Electrical constants in a transmission line are distributed in which of the following ways?

1. Into a single device
2. Along the length of the line
3. According to the thickness of the line
4. According to the cross-sectional area of the line

3-16. Leakage current in a two-wire transmission line is the current that flows through what component?

1. The resistor
2. The inductor
3. The insulator
4. The conductor

3-17. Conductance is the reciprocal of what electrical property?

1. Inductance
2. Resistance
3. Capacitance
4. Reciprocity

3-18. A transmission line that has current flowing through it has which, if any, of the following fields about it?

1. Electric field only
2. Magnetic field only
3. Both electric and magnetic fields
4. None of the above

3-19. Maximum transfer of energy from the source to the transmission line takes place when what impedance relationship exists between the source and the transmission line?

1. When the load impedance equals source impedance
2. When the load impedance is twice the source impedance
3. When the load impedance is half the source impedance
4. When the load impedance is one-fourth the source impedance

3-20. The characteristic impedance (Z_0) of a transmission line is calculated by using which of the following ratios?

1. R_s to R_{load} of the line
2. I_{max} to I_{min} at every point along the line
3. E to I at every point along the line
4. E_{in} to E_o of the line

3-21. For a given voltage, what determines the amount of current that will flow in a transmission line?

1. Conductance
2. Spacing of the wires
3. Diameter of the wires
4. Characteristic impedance

3-22. When the impedance of a transmission line is measured, which of the following values frequently is NOT considered?

1. Inductance
2. Resistance
3. Conductance
4. Capacitance

3-23. The characteristic impedance of a long transmission line may be determined by using which of the following methods?

1. Trial and error
2. Calculating the impedance of the entire line
3. Calculating the impedances at each end of the line
4. Adding the impedances of successive short sections

3-24. When should lumped values for transmission-line constants be used to calculate characteristic impedance?

1. When the line is short compared to one wavelength
2. When the line is long compared to one wavelength
3. When the line is infinitely long

3-25. In actual practice, the characteristic impedance of a transmission line is usually within which of the following resistance ranges?

1. 0 to 0.9 ohm
2. 1 to 49 ohms
3. 50 to 600 ohms
4. 601 to 1,000 ohms

3-26. The input impedance of a transmission line is affected by which of the following properties?

1. Radiation loss
2. Series inductance
3. Parallel capacitance
4. Each of the above

3-27. When a dc voltage is applied to a transmission line and the load absorbs all the energy, what is the resulting relationship between current and voltage?

1. They are in phase with each other
2. They are equal to Z_0 of the line
3. They are out of phase with each other
4. They are evenly distributed along the line

3-28. The initial waves that travel from the source to the load of a transmission line are referred to as what type of waves?

1. Incident
2. Refracted
3. Reflected
4. Diffracted

3-29. Waves that travel from the output end to the input end of a transmission line are referred to as what type of waves?

1. Incident
2. Refracted
3. Reflected
4. Diffracted

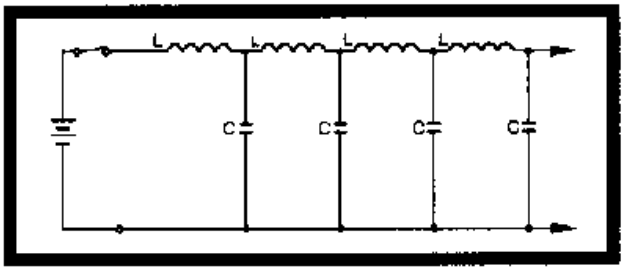

Figure 3-A.—Equivalent infinite transmission line.

IN ANSWERING QUESTION 3-30, REFER TO FIGURE 3-A.

3-30. When a dc voltage is applied to the equivalent infinite line in the figure, which of the following conditions occurs along the length of the line?

1. Standing waves of voltage form
2. Standing waves of current form
3. Current flows indefinitely
4. Voltage appears for a short time

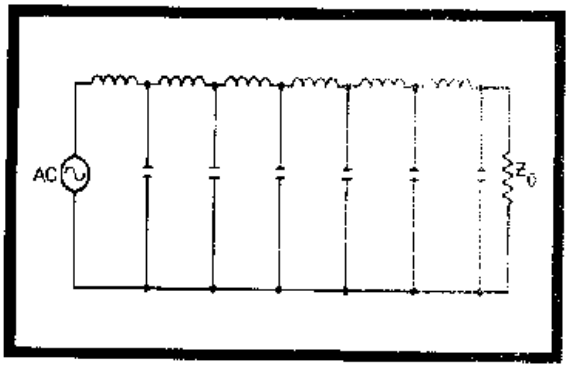

Figure 3-B.—Equivalent transmission line.

IN ANSWERING QUESTION 3-31, REFER TO FIGURE 3-B.

3-31. Compared to a dc input, what relative amount of time is required for an ac input voltage to travel the length of the line shown in the circuit?

1. Less
2. Same
3. More

3-32. The instantaneous voltage on an infinite transmission line can be plotted against time by using which of the following instruments?

1. A wavemeter
2. A multimeter
3. An oscilloscope
4. A spectrum analyzer

3-33. On an infinite transmission line with an ac voltage applied, which of the following is an accurate description of the effective voltage distribution along the line?

1. Voltage is 0 at all points
2. Voltage is constant at all points
3. Voltage varies at a sine-wave rate
4. Voltage varies at double the sine-wave rate

3-34. The velocity of propagation on a transmission line is controlled by which of the following line characteristics?

1. Conductance
2. Inductance only
3. Capacitance only
4. Capacitance and inductance

3-35. The total charge on a transmission line is equal to the current multiplied by which of the following factors?

1. Time
2. Power
3. Voltage
4. Resistance

3-36. With only capacitance and inductance of the line given, the time (T) required for a voltage change to travel down a transmission line can be found by what formula? The characteristic impedance for an infinite transmission line can be figured using which of the following ratios?

1. $T = \sqrt{\frac{L}{C}}$
3. $T = L + C$
2. $T = \sqrt{LC}$
4. $T = L - C$

3-37. The characteristic impedance for an infinite transmission line can be figured using which of the following ratios?

1. Input current to velocity
2. Input voltage to input current
3. Input voltage to line resistance
4. Input current to line resistance

3-38. The characteristic impedance of a transmission line can be figured by using which of the following formulas?

1. $Z_0 = \frac{1}{LC}$
3. $Z_0 = \frac{C}{L}$
2. $Z_0 = \sqrt{LC}$
4. $Z_0 = \sqrt{\frac{L}{C}}$

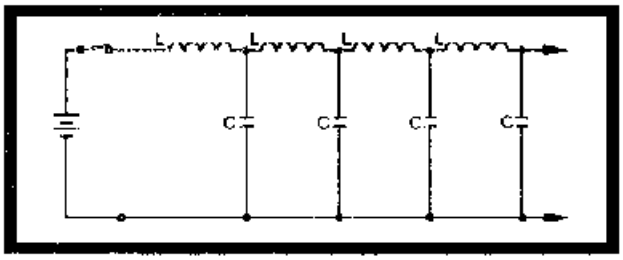

Figure 3-C.—Equivalent transmission line.

IN ANSWERING QUESTIONS 3-39 AND 3-40, REFER TO FIGURE 3-C. ASSUME THAT THE LINE IS 1,200 FEET LONG. A 150-FOOT SECTION IS MEASURED TO DETERMINE L AND C. THE 150-FOOT SECTION HAS AN INDUCTANCE OF 0.36 MILLIHENRIES AND A CAPACITANCE OF 1,000 PICOFARADS.

3-39. What is the characteristic impedance of the line?

1. 400 ohms
2. 600 ohms
3. 800 ohms
4. 900 ohms

3-40. What is the velocity of the wave on the 150-foot section?

1. 210,000,000 fps
2. 225,000,000 fps
3. 250,000,000 fps
4. 275,000,000 fps

3-41. If a transmission line is open-ended, which of the following conditions describes its terminating impedance?

1. Finite
2. Infinitely large
3. Equal to load impedance
4. Equal to source impedance

3-42. When a transmission line is not terminated in its characteristic impedance (Z_0), what happens to the incident energy that is NOT transferred to the load?

1. It is returned along the transmission line
2. It is radiated into space
3. It is absorbed by the line
4. It is converted to heat energy

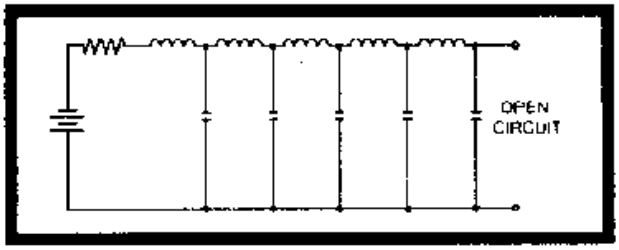

Figure 3-D.—Open-ended transmission line.

IN ANSWERING QUESTIONS 3-43 AND 3-44, REFER TO FIGURE 3-D.

3-43. When the dc voltage reaches the open end of the transmission line in the figure and is reflected, it has which, if any, of the following changes?

1. Increased amplitude
2. Decreased amplitude
3. The opposite polarity
4. None of the above

3-44. When the dc current reaches the open end of the transmission line and is reflected, it has which, if any, of the following changes?

1. Increased amplitude
2. Decreased amplitude
3. The opposite polarity
4. None of the above

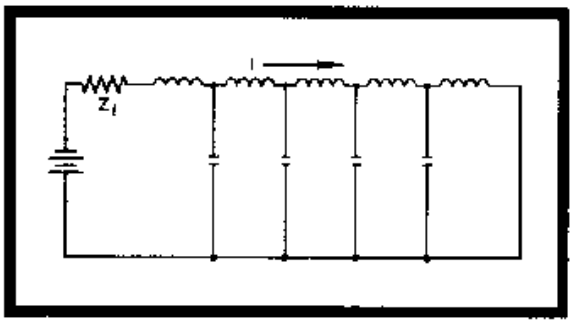

Figure 3-E.—Short-circuited transmission line.

IN ANSWERING QUESTIONS 3-45 AND 3-46, REFER TO FIGURE 3-E.

3-45. When the dc voltage reaches the shorted end of the transmission line, it is reflected. It has which, if any, of the following changes?

1. Increased amplitude
2. Decreased amplitude
3. The opposite polarity
4. None

3-46. When the dc current reaches the shorted end of the transmission line, it is reflected. It has which, if any, of the following changes?

1. Decrease amplitude
2. Increase amplitude
3. Increased polarity
4. None of the above

3-47. In an open-ended transmission line with an ac signal applied, what is the phase relationship between the incident and reflected voltage waves?

1. In phase
2. 45 degrees out of phase
3. 90 degrees out of phase
4. 180 degrees out of phase

3-48. The resultant of the incident and reflected voltage waves is called the standing wave. Its value is figured by using which of the following procedures?

1. Adding the effective values of the two waveforms
2. Algebraically adding the instantaneous values of the two waveforms
3. Algebraically subtracting the instantaneous values of the two waveforms
4. Taking the square root of the product of the incident and reflected voltages

3-49. On an open-ended transmission line that is carrying an ac signal, what is the total number of moving voltage waves?

1. One
2. Two
3. Three
4. Four

3-50. At the end of an open-ended transmission line, which, if any, of the following voltage waves is at its maximum value?

1. Incident
2. Reflected
3. Resultant
4. None

3-51. On a transmission line that is carrying an ac signal, what is the relative value of the resultant voltage wave 1/4 wavelength from the open end?

1. Maximum positive
2. Maximum negative
3. Zero

3-52. In an open-ended transmission line, the resultant ac current waveform is always zero at what point(s)?

1. At the open end only
2. 1/2 wavelength from the open-end only
3. At the open end and 1/2 wavelength from the open-end

3-53. The resultant waveform obtained by adding the incident wave to the reflected wave is referred to as a/an

1. standing wave
2. negative wave
3. algebraic wave
4. concentrated wave

3-54. On an open-ended transmission line, what is the phase relationship between the standing waves of voltage and current?

1. In phase
2. 45 degrees out of phase
3. 90 degrees out of phase
4. 180 degrees out of phase

3-55. Which of the following conditions exist at the end of a shorted transmission line?

1. Maximum voltage and minimum current
2. Maximum voltage and maximum current
3. Minimum voltage and maximum current
4. Minimum voltage and minimum current

3-56. Transmission line is considered to be nonresonant (flat) when it is terminated in which of the following ways?

1. In an impedance equal to Z_0
2. In an impedance that is infinite
3. In an inductive reactance greater than Z_0
4. In a capacitive reactance greater than Z_0

3-57. Of the following terms, which one is used for the nonresonant transmission line?

1. A tuned line
2. A shorted line
3. An untuned line
4. A terminated line

3-58. A transmission line that is resonant is sometimes referred to as which of the following types of lines?

1. Tuned
2. Matched
3. Untuned
4. Unmatched

3-59. A short-circuited section of transmission line that is an odd number of quarter-wavelengths long shows the same characteristics as which of the following devices?

1. A series-resonant circuit
2. A parallel-resonant circuit
3. An inductive reactance equal to Z_0
4. A capacitive reactance equal to Z_0

3-60. Which of the following circuits appears as a very high resistance at resonance?

1. Nonresonant
2. Series-resonant
3. Parallel-resonant
4. Each of the above

3-61. When a series-resonant circuit is resonant at a frequency above the generator frequency, it acts as what type of circuit?

1. Open
2. Resistive
3. Inductive
4. Capacitive

3-62. Which of the following sections of transmission line can be used as a parallel-resonant circuit?

1. A shorted 1/4-wavelength section
2. An open 1/4-wavelength section
3. A shorted 1/2-wavelength section
4. An open 3/4-wavelength section

3-63. A generator connected to an open-ended line greater than 1/4 wave-length but less than 1/2 wave-length senses which of the following circuit component characteristics?

1. Zero reactance
2. Low resistance
3. Inductive reactance
4. Capacitive reactance

3-64. Which of the following conditions of current (I) and impedance (Z) exist at even quarter-wave points on a shorted transmission line?

1. Low I, low Z
2. Low I, high Z
3. High I, high Z
4. High I, low Z

3-65. What is the maximum distance, in wavelengths (λ), between adjacent zero-current points on an open-circuited line?

1. 1 λ
2. 1/2 λ
3. 1/4 λ
4. 1/8 λ

3-66. When a line is terminated in a capacitance, the capacitor performs which, if any, of the following circuit actions?

1. It absorbs all the energy
2. It reflects all the energy
3. It reacts as if it were a short
4. None

3-67. When a transmission line is terminated in an inductive reactance, which, if any, of the following phase shifts takes place with respect to the current and voltage?

1. Only voltage is phase-shifted
2. Only current is phase-shifted
3. Both voltage and current are phase-shifted
4. None

3-68. When a transmission line is terminated in a resistance greater than Z_0, which of the following conditions exist?

1. The end of the line appears as an open circuit
2. Standing waves appear on the line
3. Voltage is maximum and current is minimum at the end of the line
4. Each of the above

3-69. On a transmission line, reflections begin at which of the following locations?

1. At the load
2. At the source
3. At the middle
4. At the half-wavelength point

3-70. The ratio of maximum voltage to minimum voltage on a transmission line is referred to as the

1. rswr
2. pswr
3. vswr
4. iswr

3-71. Which of the following ratios samples the magnetic field along a line?

1. Vswr
2. Pswr
3. Iswr
4. Rswr

ASSIGNMENT 4

Textbook assignment: Chapter 4, "Antennas," pages 4-1 through 4-60.

4-1. Radio energy is transmitted through which of the following mediums?

1. Rock
2. Soil
3. Water
4. Space

4-2. Energy is transmitted from a transmitter into space using which of the following devices?

1. A receiver
2. A delay line
3. A receiving antenna
4. A transmitting antenna

4-3. Transmitted rf energy takes what form as it is sent into space?

1. A magnetic field only
2. An electric field only
3. An electromagnetic field
4. A static dielectric field

4-4. The dimensions of a transmitting antenna are determined by which of the following factors?

1. Transmitted power
2. Transmitted frequency
3. Distance to the receiver
4. Antenna height above the ground

4-5. A device used to radiate or receive electromagnetic wave energy is referred to as a/an

1. feeder
2. antenna
3. transmitter
4. coupling device

4-6. An antenna that can be mounted to radiate rf energy either vertically or horizontally is classified as which of the following types?

1. Hertz
2. Marconi
3. Quarter-wave
4. Both 2 and 3 above

4-7. A complete antenna system consists of which of the following components?

1. A feeder, a coupling device, and a transmitter
2. A feeder line, a coupling device, and an antenna
3. An antenna, a transmission line, and a receiver
4. An impedance-matching device, a feeder, and a transmission line

4-8. What component in an antenna system transfers energy from the transmitter to the antenna?

1. A feeder
2. A delay line
3. A choke joint
4. A rotating joint

4-9. The type, size, and shape of an antenna are determined by which of the following factors?

1. Power output of the transmitter
2. Transmitter frequency
3. Direction to the receiver
4. Each of the above

4-10. Moving electric and magnetic fields in space have what (a) phase and (b) angular relationships?

1. (a) In phase
   (b) Perpendicular
2. (a) In phase
   (b) Displaced 45º
3. (a) Out of phase
   (b) Displaced 45º
4. (a) Out of phase
   (b) Perpendicular

4-11. What is the length of each half of the wire for a dipole antenna?

1. Wavelength
2. 3/4 wavelength
3. 1/2 wavelength
4. 1/4 wavelength

4-12. On a dipole antenna, the sinusoidal variation in charge magnitude lags the sinusoidal variation in current by what amount?

1. 1 cycle
2. 1/2 cycle
3. 1/4 cycle
4. 1/8 cycle

4-13. On a standing wave, the points of high current and voltage are identified by which of the following terms?

1. Peaks
2. Nodes
3. Poles
4. Loops

4-14. The presence of standing waves indicates which of the following conditions of an antenna?

1. Resonance
2. Saturation
3. Nonresonance
4. Minimum efficiency

4-15. The antenna property that allows the same antenna to both transmit and receive energy is

1. gain
2. resonance
3. reciprocity
4. directivity

4-16. There is a ratio between the amount of energy propagated in certain directions by a directional antenna compared to the energy that would be propagated in these directions if the antenna were not directional. This ratio is known as which of the following antenna characteristics?

1. Gain
2. Directivity
3. Reciprocity
4. Polarization

4-17. The polarization plane of the radiation field is determined by which of the following fields?

1. Electric field only
2. Magnetic field only
3. Electromagnetic field

4-18. For best reception of a signal from a horizontally polarized antenna, the receiving antenna should be mounted so that it has what relationship with the transmitting antenna?

1. 0 degrees
2. 45 degrees
3. 90 degrees
4. 135 degrees

4-19. An electric field that rotates as it travels through space exhibits what type of polarization?

1. Vertical
2. Spherical
3. Elliptical
4. Horizontal

4-20. For ground-wave transmissions, what type of polarization is required?

1. Vertical
2. Spherical
3. Elliptical
4. Horizontal

4-21. For high-frequency operation, which of the following antenna polarization patterns is preferred?

1. Vertically polarized
2. Spherically polarized
3. Elliptically polarized
4. Horizontally polarized

4-22. Omnidirectional transmission is obtained from which of the following antennas?

1. Elliptically polarized
2. Horizontal half-wave
3. Vertical half-wave
4. Each of the above

4-23. With an antenna height of 40 feet and a transmitter frequency of 90 megahertz, which of the following antenna radiation patterns is best for transmitting over bodies of water?

1. Vertically polarized
2. Spherically polarized
3. Elliptically polarized
4. Horizontally polarized

4-24. To select a desired signal and discriminate against interfering signals from strong vhf and uhf broadcast transmissions, which of the following actions should you take?

1. Increase receiver gain
2. Make the transmitting antenna bi-directional
3. Use a vertically polarized receiving antenna
4. Use narrowly directional arrays as receiving antennas

4-25. A vertically mounted transmission line is LEAST affected by which of the following antenna radiation patterns?

1. Vertically polarized
2. Spherically polarized
3. Horizontally polarized
4. Elliptically polarized

4-26. An antenna with which of the following radiation resistance values will exhibit reduced efficiency?

1. 39 ohms
2. 82 ohms
3. 107 ohms
4. 150 ohms

4-27. An isotropic radiator radiates energy in which of the following patterns?

1. Vertical
2. Bi-directional
3. Unidirectional
4. Omnidirectional

4-28. An ordinary flashlight is an example of what type of radiator?

1. Isotropic
2. Polarized
3. Anisotropic
4. Stroboscopic

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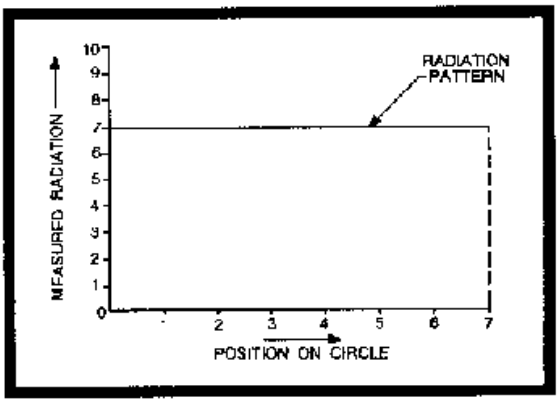

Figure 4-A.—Rectangular-coordinate graph.

IN ANSWERING QUESTION 4-29, REFER TO FIGURE 4-A.

4-29. How many points on the graph can represent the value of 7 radiation units at position 2 of the circle?

1. One
2. Two
3. Three
4. Four

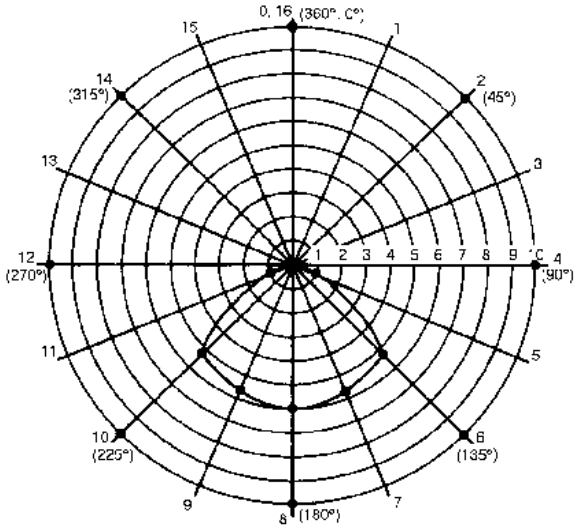

Figure 4-B.—Polar-coordinate graph.

IN ANSWERING QUESTIONS 4-30 AND 4-31, REFER TO FIGURE 4-B.

4-30. Compared with the rectangular-coordinate graph, the polar-coordinate graph has the advantage of showing which of the following antenna characteristics?

1. Polarization
2. Radiation pattern
3. Phase relationship
4. Gain versus directivity

4-31. The area enclosed by the radiation pattern is the

1. lobe
2. null
3. axis
4. coordinate

4-32. Inserting an inductor or capacitor in series with an antenna is one method of electrically changing the electrical length of an antenna. What is this method called?

1. Loading
2. Inserting
3. Unloading
4. Decoupling

4-33. Many complex antennas are constructed from what basic antenna?

1. The Marconi antenna
2. The full-wave antenna
3. The half-wave antenna
4. The quarter-wave antenna

4-34. On an energized half-wave antenna, which of the following electrical conditions exist?

1. Voltage is maximum at the ends
2. Voltage is minimum at the ends
3. Current is maximum at the ends
4. Impedance is minimum at the center

4-35. Which of the following radiation patterns is/are exhibited by a simple vertical doublet antenna?

1. Nondirectional in the horizontal plane
2. Directional in the vertical plane
3. Both 1 and 2 above
4. Spherical in all planes

4-36. A method of feeding energy to a half-wave antenna is to connect one end through a capacitor to the output stage. What is this method of feeding called?

1. End feed
2. Voltage feed
3. Both 1 and 2 above
4. Current feed

4-37. An antenna supplied by the center-feed method is fed at what point?

1. Low voltage and low current
2. Low voltage and high current
3. High voltage and low current
4. High voltage and high current

4-38. The basic Marconi antenna has which of the following characteristics?

1. One-quarter wavelength and ungrounded
2. One-half wavelength and grounded at one end
3. One-half wavelength and insulated from ground
4. One-quarter wavelength and grounded at one end

4-39. The Marconi antenna behaves as a dipole for which of the following reasons?

1. It is fed at one end
2. An image antenna is formed by reflections from the ground
3. A quarter-wavelength of conductor is buried in the ground and forms the rest of the dipole
4. The applied signal is rectified so that only half the signal will appear on the quarter-wave antenna

4-40. A series of conductors arranged in a radial pattern and buried in the ground beneath the antenna is referred to as a

1. ground spur
2. counterpoise
3. ground screen
4. ground reflector

4-41. A folded dipole can be used instead of a simple, center-fed dipole for which of the following purposes?

1. Matching voltage
2. Matching impedance
3. Increasing directivity
4. Decreasing directivity

4-42. An antenna arrangement that has elements aligned in a straight line is referred to as what type array?

1. Isotropic
2. Collinear
3. Line-of-sight
4. Unidirectional

4-43. To have current in two adjoining collinear half-wave elements in proper phase, they must be connected by which of the following stubs?

1. A shorted half-wave stub
2. An open quarter-wave stub
3. A shorted eighth-wave stub
4. A shorted quarter-wave stub

4-44. To select a desired signal and discriminate against interfering signals, the receiving antenna should have which of the following characteristics?

1. Be omnidirectional
2. Be highly directional
3. Be vertically polarized
4. Be horizontally polarized

4-45. Adding more elements to a collinear antenna array produces which of the following effects?

1. Increased gain
2. Decreased gain
3. Decreased directivity
4. Mismatched impedances

4-46. What is the maximum number of elements ordinarily used in a collinear array?

1. One
2. Two
3. Three
4. Four

4-47. Constructing a collinear array with elements longer than 1/2 wavelength has which of the following effects on antenna characteristics?

1. Increased gain
2. Decreased gain
3. Increased frequency range
4. Decreased frequency range

4-48. In a two-element collinear array, maximum gain is obtained when the center-to-center spacing between the ends of the elements is approximately what electrical distance?

1. Wavelength
2. 0.15 wavelength
3. 0.5 wavelength
4. 0.75 wavelength

4-49. Compared with collinear arrays, broadside arrays have which of the following advantages?

1. Sharper tuning
2. Broader bandwidth
3. Broader frequency response
4. Less coupling between dipole

4-50. Optimum gain is obtained from a broadside array when the spacing of its elements is which of the following distances?

1. One-half wavelength
2. One-quarter wavelength
3. Greater than one-half wavelength
4. Slightly less than one-quarter

4-51. An end-fire array physically resembles the collinear array except that it is more compact. What disadvantage does the endfire array possess?

1. It has lower gain
2. It has low radiation resistance
3. It has loose coupling
4. Each of the above

4-52. What is the range of electrical spacing between the elements of an end-fire array?

1. 3/4 to 1 wavelength
2. 1/2 to 3/4 wavelength
3. 1/4 to 1/2 wavelength
4. 1/8 to 1/4 wavelength

4-53. The end-fire array produces what type of lobes, if any, along the axis of the array?

1. Minor lobes
2. Major lobes
3. None

4-54. Assuming that the elements are correctly spaced, the directivity of an end-fire array may be increased by which of the following actions?

1. Increasing the frequency
2. Decreasing the frequency
3. Decreasing the number of elements
4. Increasing the number of elements

4-55. A unidirectional pattern can be obtained from an end-fire array by using what phase relationship between the energy fed to adjacent elements?

1. 0º
2. 45º
3. 90º
4. 180º

4-56. Energy is fed to a parasitic element using what method?

1. Direct coupling
2. Inductive coupling
3. Capacitive coupling
4. Transmission-line coupling

4-57. The directivity pattern resulting from the action of parasitic elements depends on which of the following element characteristics?

1. Length of the element
2. Diameter of the element
3. Spacing between parasitic and driven elements
4. Each of the above

4-58. The advantages of unidirectivity and increased gain can best be obtained by using which of the following elements in a parasitic array?

1. Driven elements only
2. Reflector and director elements only
3. Reflector, director, and driven elements
4. Driven and director elements only

4-59. The ratio of energy radiated by an array in the principal direction of radiation to the energy radiated in the opposite direction describes which of the following relationships?

1. Side-to-side ratio
2. Front-to-back ratio
3. Driven-to-parasitic ratio
4. Reflector-to-director ratio

4-60. The Yagi antenna is an example of what type of antenna array?

1. Driven
2. End-fire
3. Multielement parasitic
4. Single-element parasitic

4-61. The addition of parasitic elements to the Yagi antenna has which of the following effects on antenna characteristics?

1. Increased gain
2. Narrower beam width
3. Narrower frequency response
4. Each of the above

4-62. An antenna which is designed especially for vertically-polarized ground waves at low frequencies is the

1. Yagi antenna
2. Marconi antenna
3. Beverage antenna
4. V antenna

4-63. What is the phase relationship of the signals that feed the V antenna?

1. 0º
2. 45º
3. 90º
4. 180º

4-64. A rhombic antenna is essentially a combination of which of the following antennas?

1. Two stacked long-wire radiators
2. Two V antennas placed side by side
3. Two collinear arrays in parallel
4. Four parallel half-wave radiators

4-65. A rhombic antenna has which of the following advantages?

1. Simple construction
2. Wide frequency range
3. Noncritical adjustment
4. Each of the above

4-66. The principal disadvantage of the rhombic antenna is its

1. poor directivity
2. large antenna site
3. low antenna voltage
4. high-frequency inefficiency

4-67. The unidirectional radiation pattern of the rhombic antenna is caused by which of the following antenna characteristics?

1. Size
2. Shape
3. Termination resistance
4. Frequency of the input energy

4-68. Horizontal half-wave antennas mounted at right angles to each other in the same horizontal plane make up which of the following antennas?

1. Rhombic
2. Flat-top
3. Turnstile
4. Ground-plane

4-69. The most common means of obtaining a low-radiation angle from a vertical quarter-wave antenna is by what procedure?

1. Decreasing power
2. Increasing frequency
3. Adding a ground plane
4. Rotating the antenna to a horizontal plane

4-70. A corner reflector antenna is used for which of the following purposes?

1. To decrease frequency range
2. To increase frequency range
3. To produce a unidirectional pattern
4. To produce an omnidirectional pattern

4-71. If a corner-reflector antenna is horizontally polarized, its radiation pattern will take on what shape?

1. A narrow beam in the horizontal plane
2. A narrow beam in the vertical plane
3. A beam similar to a half-wave dipole in the horizontal plane
4. A beam similar to a half-wave dipole with a reflector in the vertical plane

4-72. When radio or radar antennas are energized by transmitters, you must not go aloft until which of the following requirements are met?

1. A safety harness has been issued to you
2. All transmitters are secured and tagged
3. A working aloft "chit" has been filled out and signed by proper authority
4. Each of the above

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ANTENNA BASICS

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What are the basics of antennas?

Antennas, to quote a friend, are one of life's eternal mysteries. "All I'm totally certain of is that any antenna is better than no antenna and the antenna should preferably erected as high and be as long as is possible or desirable". Here we will discuss the very basics of antennas. Remember that thought: **these are just some introductory antenna basics**. Each type of antenna will eventually have its own page. In particular I would commend everyone to read my page on earth dangers. I think it ought to be compulsory reading.

The basic antenna

The most basic antenna is called "a quarter wave vertical", it is a quarter wavelength long and is a vertical radiator. Typical examples of this type would be seen installed on motor vehicles for two way communications. Technically the most basic antenna is an "isotropic radiator". This is a mythical antenna which radiates in all directions as does the light from a lamp bulb. It is the standard against which we sometimes compare other antennas.

This type of antenna relies upon an "artificial ground" of either drooping radials or a car body to act as ground. Sometimes the

antenna is worked against an actual ground - see later.

Antenna Polarisation

Depending upon how the antenna is orientated physically determines it's polarisation. An antenna erected vertically is said to be "vertically polarised" while an antenna erected horizontally is said (not so surprising) to be "horizontally polarised". Other specialised antennas exist with "cross polarisation", having both vertical and horizontal components and we can have "circular polarisation".

Note that when a signal is transmitted at one polarisation but received at a different polarisation there exists a great many decibels of loss.

This is quite significant and is often taken advantage of when TV channels and other services are allocated. If there is a chance of co-channel interference then the license will stipulate a different polarisation. Have you ever noticed vertical and horizontal TV antennas in some areas. Now you know why.

Antenna Impedance

Technically, antenna impedance is the ratio at any given point in the antenna of voltage to current at that point. Depending upon height above ground, the influence of surrounding objects and other factors, our quarter wave antenna with a near perfect ground exhibits a nominal input impedance of around 36 ohms. A half wave dipole antenna is nominally 75 ohms while a half wave folded dipole antenna is nominally 300 ohms. The two previous examples indicate why we have 75 ohm coaxial cable and 300 ohm ribbon line for TV antennas.

A quarter wave antenna with drooping quarter wave radials exhibits a nominal 50 ohms impedance, one reason for the existence of 50 ohm coaxial cable.

The quarter wave vertical antenna

The quarter wave vertical antenna is usually the simplest to construct and erect although I know a great many people who would dispute that statement. In this context I am speaking of people (the majority) who have limited space to erect an antenna.

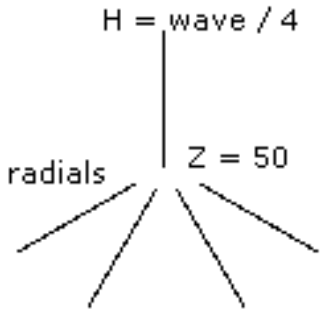


Figure 1. - a quarter wave vertical antenna with drooping radials

In figure 1 we have depicted a quarter wave vertical antenna with drooping radials which would be about 45 degrees from horizontal. These 45 degree drooping radials simulate an artificial ground and lead to an antenna impedance of about 50 ohms.

A quarter wave vertical antenna could also be erected directly on the ground and indeed many AM radio transmitting towers accomplish this especially where there is suitable marshy ground noted for good conductivity. An AM radio transmitting tower of a quarter wave length erected for say 810 Khz in the AM band would have a length of nearly 88 metres (288') in height.

The formula for quarter wave is L = 71.25 metres / freq (mhz) and in feet L = 234 / freq (mhz). Note the variance from the standard wavelength formula of 300 / freq. This is because we allow for "velocity factor" of 5% and our wavelength formula becomes 285 / freq.

When a quarter wave antenna is erected and "worked" against a good rf ground (called a Marconi Antenna) the earth provides a "mirror" image of the missing half of the desired half wave antenna.

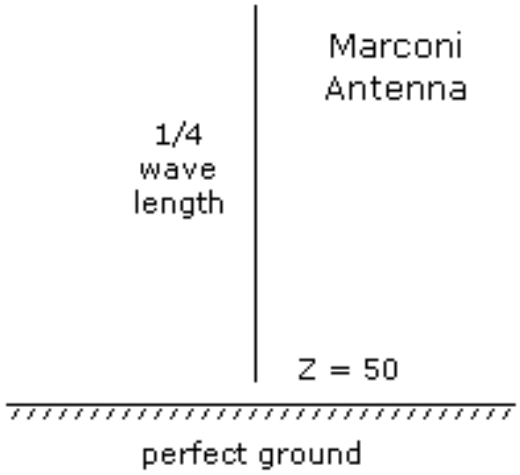


Figure 2. - a marconi antenna

In figure 2 above where I have depicted the Marconi Antenna imagine a duplicate of the quarter wave antenna being in existence from the top of the ground and extending down the page. This is the mirror image.

Half wave dipole antenna

The half wave dipole antenna becomes quite common where space permits. It can be erected vertically but is more often than not erected horizontally for practical reasons. I gave quite a good example of its use in my paper on radio telescopes from my original site. I have reproduced it in figure 3 below.

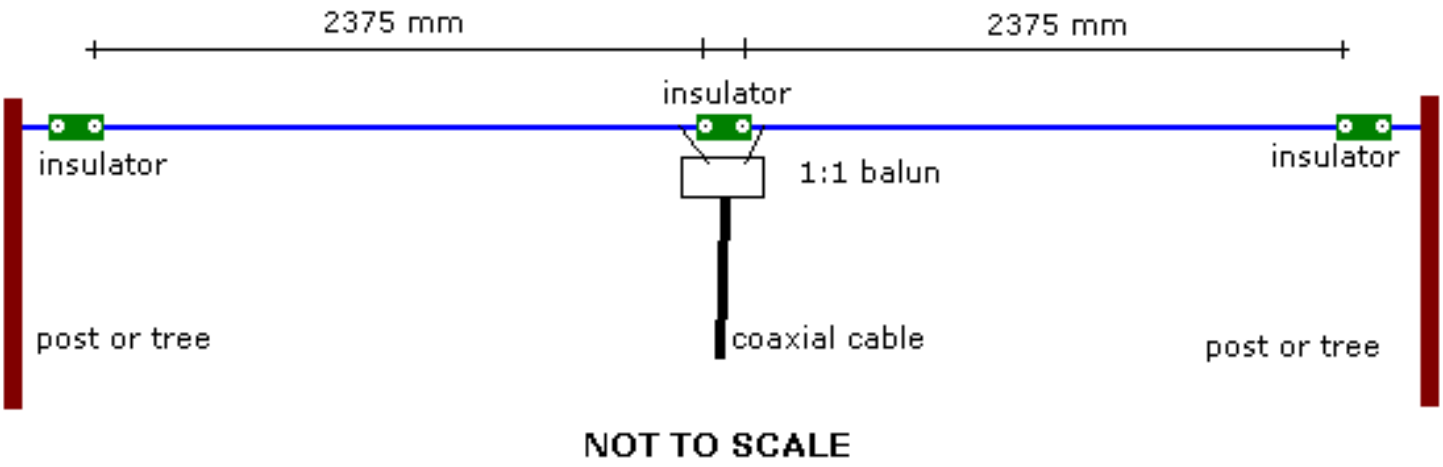


Figure 3. - half wave dipole antenna

This particular antenna was dimensioned for use at 30 Mhz. You will note that the left and right hand halves are merely quarter wave sections determined by the formula given earlier. The input impedance (affected by many factors) is nominally 50 ohms.

As with all antennas, the height above ground and proximity to other objects such as buildings, trees, guttering etc. play an important part. However, reality says we must live with what we can achieve in the real world notwithstanding what theory may say.

People erect half wave dipoles in attics constructed of fine gauge wire - far from ideal BUT they get reasonable results by living with less than the "ideal". A lesson in life we should always remember in more ways than one.

The folded dipole antenna

The folded dipole antenna is probably only ever seen as a TV antenna. It exhibits an impedance of 300 ohms whereas a half wave dipole is 75 ohms and I'm certain someone will be alert enough to ask "why 75 ohms, if figure 3 above is 50 ohms?".

Within the limits of my artistic skills I have depicted a folded dipole antenna below.

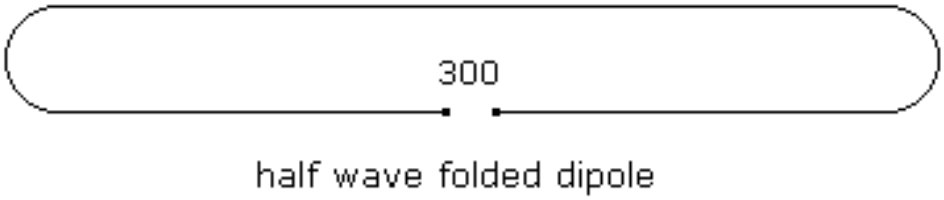


Figure 4. - half wave folded dipole

One powerful advantage of a folded dipole antenna is that is has a wide bandwidth, in fact a one octave bandwidth. This is the reason it was often used as a TV antenna for multi channel use. Folded dipole antennas were mainly used in conjuction with Yagi antennas.

The Yagi antenna

The Yagi antenna or more correctly, the Yagi - Uda antenna was developed by Japanese scientists in the 1930's. It consists of a half wave dipole (sometimes a folded one, sometimes not), a rear "reflector" and may or may not have one or more forward "directors". These are collectively referred to as the "elements".

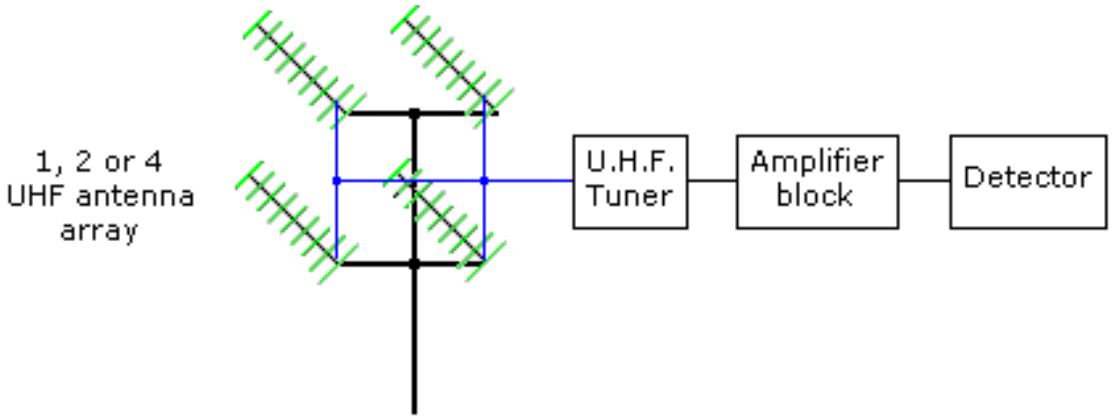


Figure 5. - the Yagi antenna

In figure 5 above I have reprinted a UHF Yagi antenna array from my radio telescopes page. You will note, not altogther clearly.

However in figure 6 below, which happens to be a photograph of a neighbour's TV antenna, I can clearly point out details of a practical Yagi antenna.

This particular antenna has been optimised for dual band operation. It is designed to pick up both VHF and UHF transmissions. Because I live in a regional of NSW in Australia, TV antennas tend to be single channel types designed either for higher gain or better directivity. Different examples will be presented later.

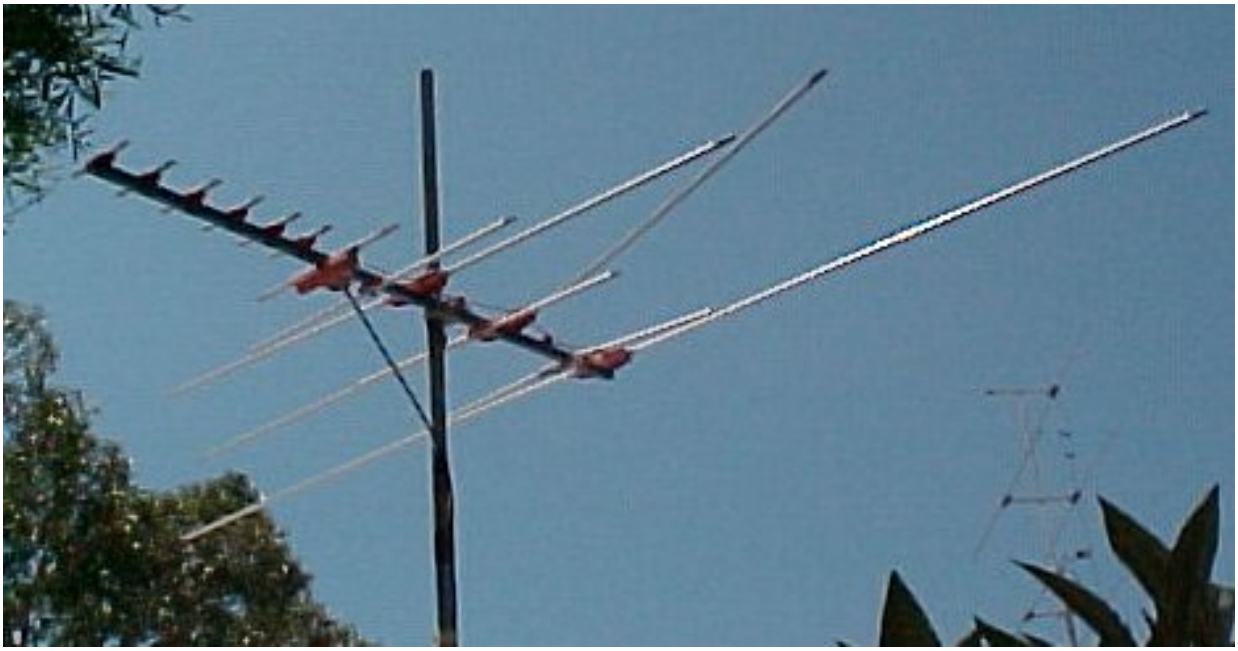


Figure 6. - a practical Yagi TV antenna

Looking from left to right on this dual band Yagi we have six UHF "director" elements which improve gain and directivity. Next is the UHF half wave dipole which could have easily been a folded dipole but is in fact a plain half wave dipole.

The next three much longer elements form a "phased array" for the

VHF band. I am unsure of the function of the three remaining smaller elements, information is quite scant here but one would certainly be a UHF "reflector". Likely the other two also fulfill this function also.

Note: This is a horizontally polarised antenna and is orientated roughly NNW, 315 degrees.

You will notice the effect of very strong storms from the sea have had in bending the second larger elements. In my locality storms are a problem but not as much as roosting parrots such as large sulphur crested cockatoos.

UHF Yagi antenna

In the photograph in figure 7 below you can see a classic UHF Yagi antenna. It has a total of nineteen "elements" comprising seventeen "directors", a fancy folded dipole with a "low-noise mast head amplifier" and a "reflector".

Figure 7. - a vertically polarised UHF Yagi antenna

This is a a vertically polarised UHF Yagi antenna and it is orientated WSW or 225 degrees. It does in fact pick up signals about 100 Km or 60 mile distant from Sydney.

This is the very same antenna I was suggesting to be used in the radio telescope array I depicted in figure 5 above.

Stacked half wave dipoles or a collinear array

The majority of TV antennas in my retirement village are stacked half wave dipoles. These consist of four sets of a half wave dipole and a reflector only, but mounted one above another. These antennas owe their origin to the days we only had VHF TV in the area. Surprising with the introduction of UHF they continued to function quite well in picking up UHF as well. This particular antenna is my one and I've never had the need to go to a UHF antenna. The top two elements normally are home to roosting "top knot" pidgeons, a pigeon native to Australia.

Figure 8. - four stacked half wave dipoles collinear antenna

To the left of the photograph are the "reflectors" and to the right are the four vertically stacked half wave dipoles. The wires connecting each half wave dipole are done in a "phased way". This comprises a collinear antenna array and is so arranged for improved gain.

Note this antenna is horizontally polarised.

Loop Antennas

The loop antenna comes in an amazing number of configurations. It is a "small space" antenna and although extremely inefficient is capable of surprising results. In receiving applications the loop antenna works on the principle of the **"differences"** in voltages induced by the current flowing in the sides of the antenna. As you might imagine these difference voltages can be extremely minute in amplitude and any loop antenna usually requires an associated amplifier capable of at least 25 dB power gain following it.

One example of a shielded loop antenna is taken from my tutorial on mobius winding techniques is shown in figure 9 below.

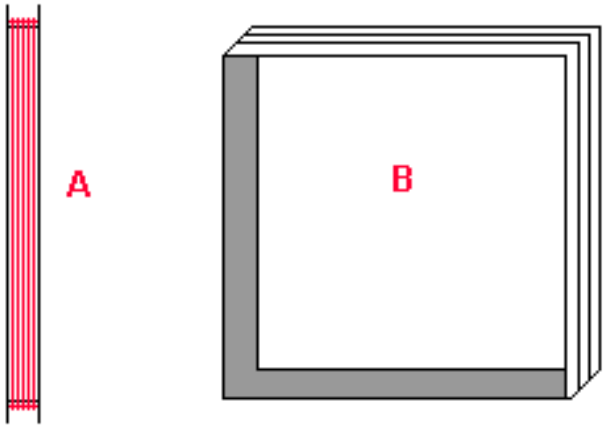


Figure 9. - mobius winding of a loop antenna

This is the general loop antenna which has one interesting characteristic. It responds well to signals arriving in one direction, either from the left hand side of your computer screen or the right hand side of your computer screen for the loop shown in figure 9 (b) above. Signals from either your face or from behind your monitor would produce equal signal currents from both sides of the loop and consequently produce no difference voltage output.

Technically speaking, a loop antenna responds to the magnetic field

rather than the electric field.

Rather than being omnidirectional (as a whip antenna would be) the loop antenna responds to the cosine of the angle between its face and the direction of arrival of the electromagnetic wave. This actually produces a figure eight pattern, which for receiving presents no probems. The addition of a small whip antenna in conjuction with proper phasing allows the direction ambiguity to be resolved and we have an antenna relatively ideal for direction finding.

The most common loop antenna you will encounter is the loopstick antenna [in the U.K. it is referred to as a "ferrite rod antenna"] built into portable receivers. In figure 10 below is the AM and shortwave loopstick antenna in a Sanyo model RP2127 MW / SW receiver (it's old).

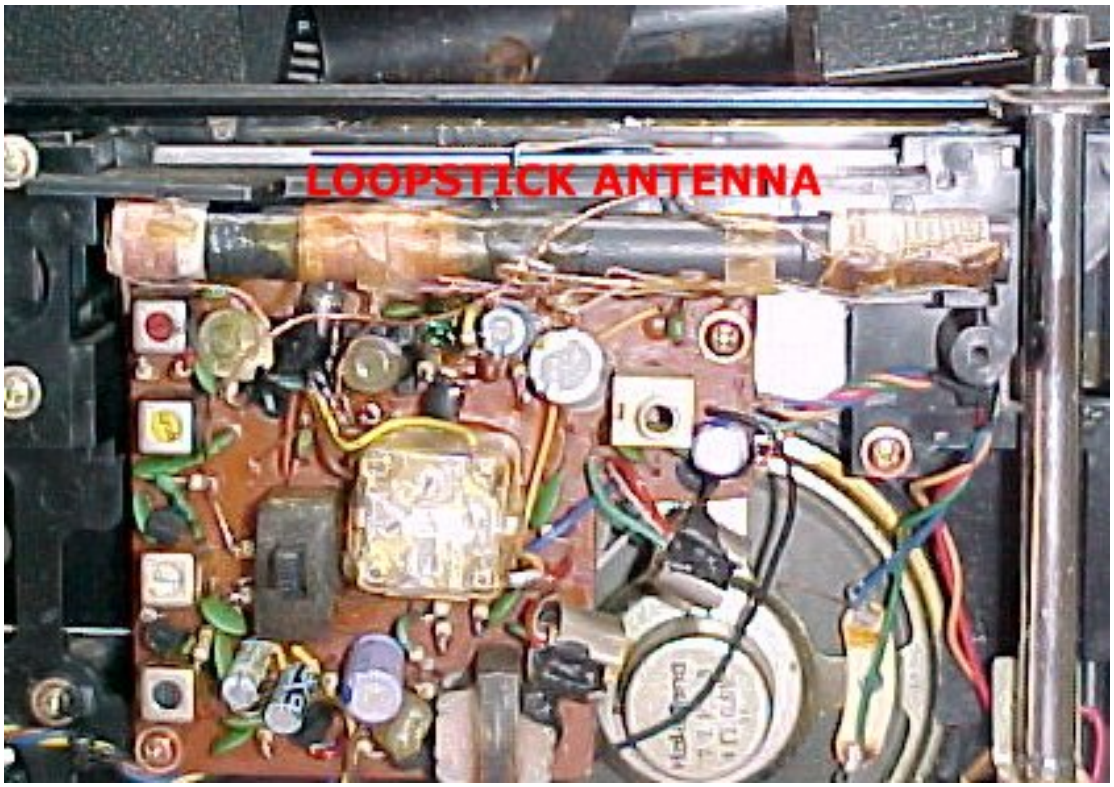


Figure 10. - AM and shortwave loopstick antenna

The AM and shortwave loopstick antenna is located in the upper half under the words "loopstick antenna". For greater efficiency and size reduction, a loopstick antenna is wound on a "ferrite" rod. This

particular one happens to be circular but you may encounter ones which are rectangular.

As an experiment you might, if you have a loopstick antenna radio available, tune to a weak station and rotate the radio around 360 degrees. You should notice two points 180 degrees apart where the signals seem to be the strongest and similarly notice two other points 180 degrees apart where the signals seem to be the weakest - these are called "nulls". This is the aid to "Radio Direction Finding - RDF"

Terminated Tilted Folded Dipole

Now here is a little gem. The terminated tilted folded dipole is bound to give a "rush of blood to the head" of any avid DX'er (that means long distance -dx- receive / transmit enthusiast).

The terminated tilted folded dipole is somewhat similar to the half wave folded dipole in figure 4 above yet the claims for its performance are quite astonishing. The terminated tilted folded dipole is claimed to have a bandwidth of something like 5 or 6 to one, been extensively tested and adopted by the US Navy, easy to construct from readily available materials and, has a feedpoint impedance of around 300 ohms.

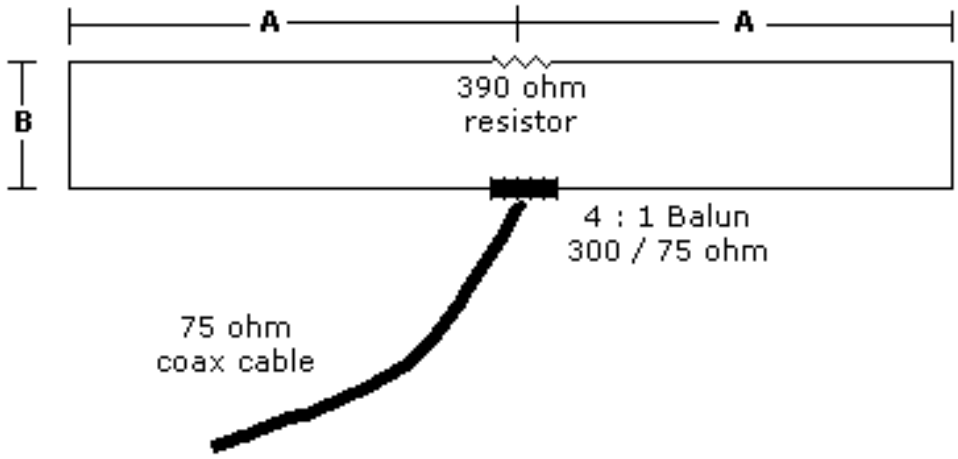


Figure 11. - Terminated Tilted Folded Dipole

The dimensions "A" and "B" for a terminated tilted folded dipole are as follows:

Each leg "A" = [2 X pi (15.25 / Fo)] and;

Distance "B" = [2 X pi (0.915 / Fo)]

where in both instances 2 X pi = 6.28 and Fo is in Mhz.

There seems to be some debate about the exact formula, my friend L. B. Cebik (see next) says:

"The "Wide-Long" version coincides with standard construction formulations, since the antenna is about 300/F(MHz) long and 10/F(MHz) wide. (Excessively fussy cutting formulas for this antenna are largely superfluous, since strict resonance is not in question)."

My friend L. B. Cebik (see later) has modeled this antenna. **Modeling the T2FD**

Further comprehensive details on the claims for the amazing terminated tilted folded dipole antenna and its construction can be found at:
http://www.hard-core-dx.com/nordicdx/antenna/wire/t2fd.html

Conclusion on antenna basics

The reason there has been emphasis on TV antennas is simply because nearly everyone can look at examples in their own locality for comparison. At TV frequencies the physical dimensions are such I can offer practical examples with photographs.

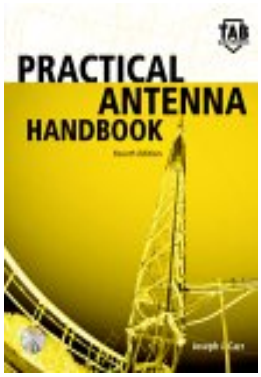
The same basic principles apply at HF and LF although physical sizes tend to be totally impractical.

As time permits I will flesh out more and more in depth articles on all these antennas and even more types not even mentioned here. This page alone comprises well over 2,000 words so you can imagine the job ahead with competing demands on my time. Meanwhile consider this important publication on antennas.

Meanwhile I would also suggest that you take a good look at L.B. Cebik 's W4RNL great web site. My good friend LB is "THE antenna guru".

PRACTICAL ANTENNA HANDBOOK

By Joseph Carr, from TAB Books



Designed for use by the novice as well as the professional, this book/CD- ROM combo gives the reader all kinds of projects with material that explains why they work.

A wide variety of antennae are covered: high frequency dipole, vertically polarized HF, multiband and tunable wire, hidden and limited space, directional phased vertical and directional beam, VHF/UHF transmitting and receiving, shortwave reception, microwave, mobile, marine and emergency.

This third edition has new material on wire antenna construction methods, antenna modeling software, antennas for radio astronomy and Radio Direction Finding, and antenna noise temperature.

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This is a mutual help group with a very professional air about it. I've learn't things. Not for absolute newcomers to ask questions BUT it is an excellent learning resource for lurkers.

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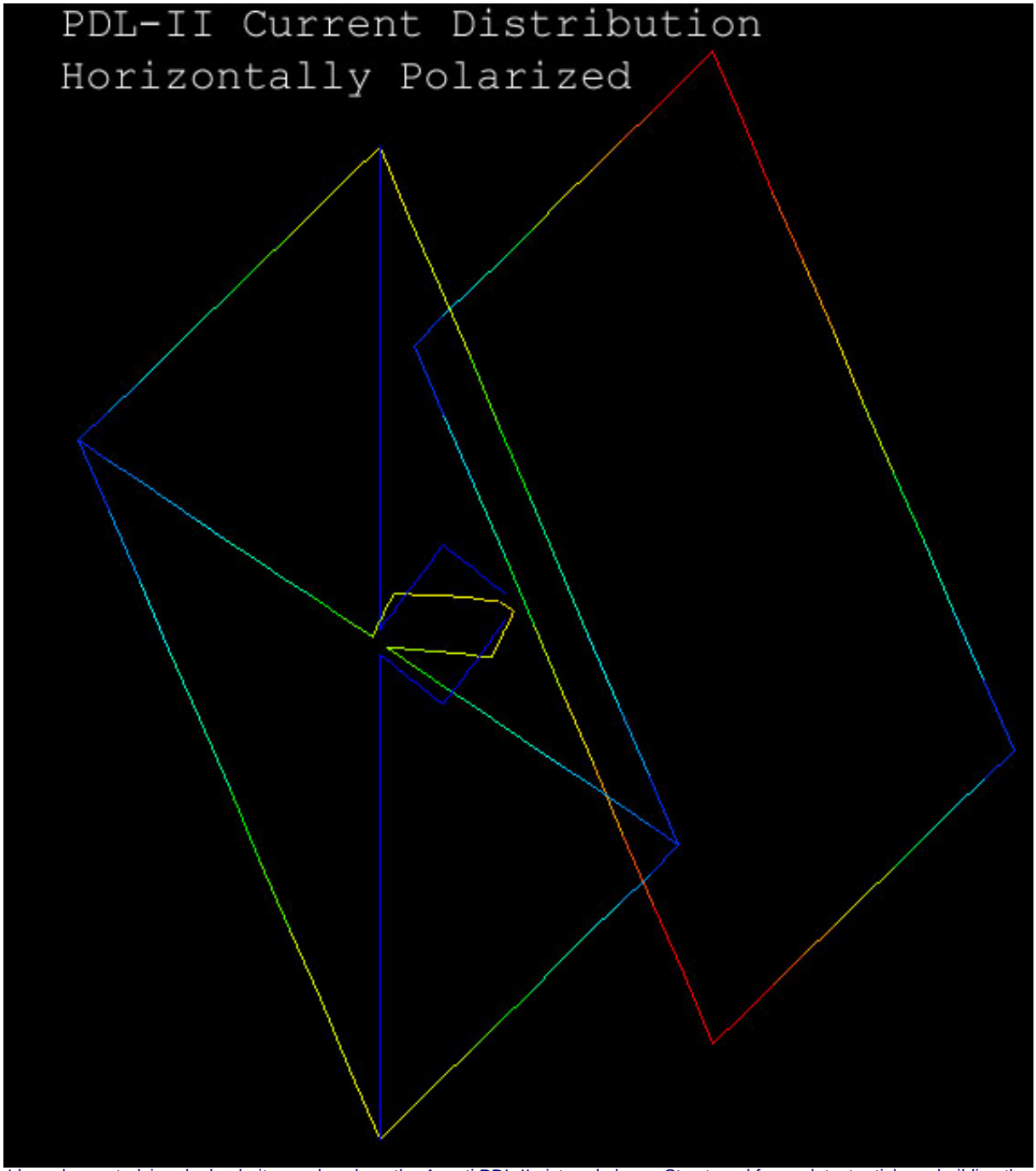

I have been studying dual polarity quad such as the Avanti PDL-II pictured above. Stay tuned for my latest article on building the ultimate dual polarity cubical quad!

Last update: Wednesday, 19-Nov-2003 23:40:43 EST Updated!

Hello, this site is dedicated to bringing true information about the different types of antennas you can use, buy and build for Citizen Band Radio and 11 Meter DXing. From concept to construction, I have set out to provide the most technically correct information possible. It contains information from antenna basics to advanced antenna theories such as the angle of radiation. I believe that it is not necessary to have great knowledge of electric fundamentals to understand how an antenna works. I have included tons of figures so that you can see what I am talking about! This site should be of interest to all CB users and 11 meter DXers because the station antenna provides the most improvement in station performance (even more so than an amplifier) when it is built correctly.

I decided to do this after all the info I read at antenna manufactures web sites. They all try to sell you their antenna. On one site I read says top loaded mobile antennas are the best, even better than base loaded. Then you read the web site of a company that makes base load mobiles only to find out base loaded are the best. Good thing is, I am not trying to sell you an antenna. Most of these web pages (and ad's in magazines) are done by the advertising department, not by the antenna engineers. And those advertised gain numbers are not regulated by any government agency! If you are looking for quick answers, frequently asked questions, this is not the place to start reading. I recommend reading it all, so that you get all the important concepts. Another reason I decided to this page is lack of good information on the web about antennas in general. Everyone has a "resource" site with a 1000 dead links to antennas manufacture pages....but no one has any *real* content. So here is my attempt at it! If there are any section that you would like to see added, let me know. For further reading see my BIBLIOGRAPHY

This page *has* to be viewed at a resolution *greater* than 800x600. I am sorry if you can not turn up the resolution any higher than 800x600, but I had to make the figures big enough to show detail clearly. You know your resolution is too small if you have to scroll out to read the end of each line in the sections (not this page, however). Click HERE for instructions on how to turn the resolution up in Windows 95/98/NT.

The Yagi building page is up!

- Scott, 2 Sugar Delta 789

Antenna Basics

Where to start reading if you are not really familiar with antenna basics such as radiation resistance, SWR, antenna bandwidth and polarization. Even if you have heard of these terms, read on to get a better understanding of all the essential fundamental antenna subjects. Discussion on the different type of antennas will use terminology explained here.

Coax Basics

Information about coaxial cable. Wonder why changing the your coax length changes the SWR of your antenna? Subjects such as impedance and velocity factor will be explained.

Antenna Types

Details the most common types of antennas used on 11 meters. Different types of omnidirectional and beams are covered and explains which design most common commercial CB antennas use. This is the section to read if you do not want to build your own antenna but figure out which commercial design best suites you. Make sure you read the "Hybrid" section if you own a Moonraker (or any beam with a quad reflector) for an important note.

Mobile Antennas

Explains mobile antenna basics.

Performance Tips

Four ways to improve your stations performance greatly.

- Co-Phase your antennas for 3db more gain
- Circular Polarize your dual polarity beam for solid DX contacts
- Angle of Radiation, Match the height of your antenna to the distance you are trying to communicate
- Really Ground that radio for improved receiver performance!

Antenna Building

Build your own optimized DX antenna. Not for the faint of heart. Now includes a JavaScript Calculator to fully automate designing the antenna for the frequency you talk on (Cubical Quad)!

- 4 Element Cubical Quad
- 4 Element Yagi NEW!

Scott <scott2RP789@netscape.net>

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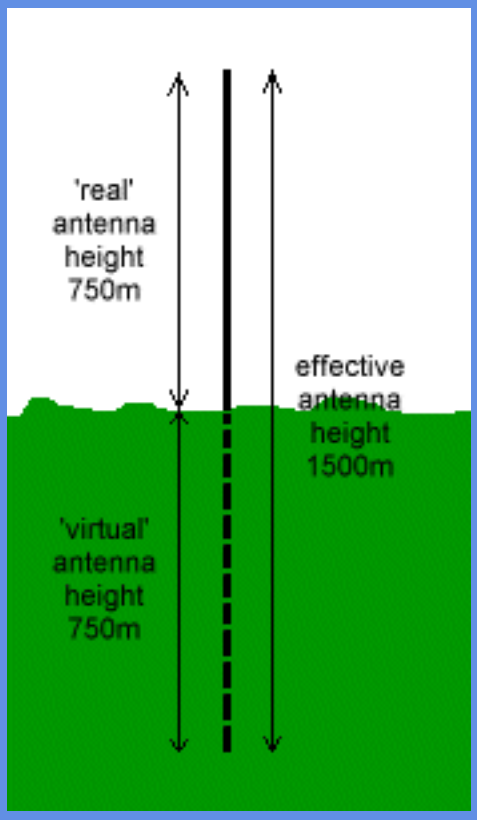
Basic Antenna Theory

Radio waves are generated by electrons accelerating in the antenna.

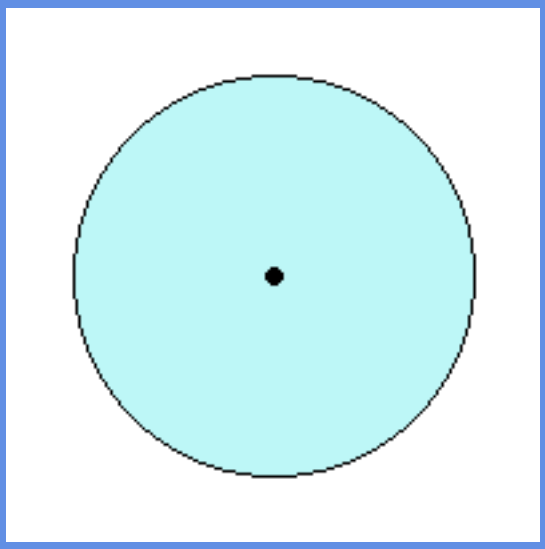
Consider a transmitter perpendicular to the ground. The electrons in the antenna, when a signal is applied, are changing their velocities continuously (i.e. moving up and down very quickly) in response to the applied signal.

For a station that broadcasts at a wavelength of 1500m, the antenna needs to be 750m long. This is because there is a 'virtual antenna' caused by the aerial being earthed in the ground:



The transmitting aerial (and the receiving aerial) need only be half-the-wavelength tall.

Now if this transmitter has no directional properties (i.e. it radiates in all directions equally), it has a coverage area, assuming completely flat ground, that is a perfect circle:



(View from above - antenna in centre; blue is coverage area)

Broadcasters rarely use a non-directional aerial though. It is possible to 'force' the energy radiated by the transmitter into particular directions - the aerial becomes directional. Directional aerials are used to great effect near the coasts of the UK, where the broadcasters do not want their signal to be easily picked up on the continent.

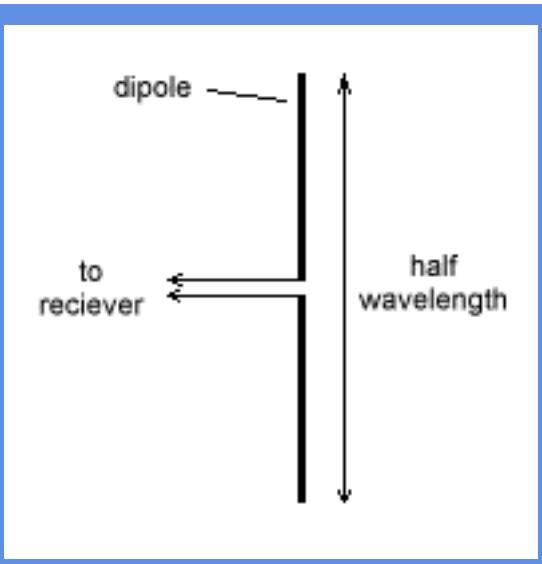
It is important that signal doesn't leak onto the continent since i) continental stations use the same frequencies and leaking signal would cause interference and ii) some programme broadcast rights apply to the UK only.

Under unusual weather conditions, despite the best efforts of broadcasters both in this country and abroad, signals travel much further than they normally would and interfere with reception of stations using the same channel, causing co-channel intereference.

Let's switch the emphasis from the transmitting aerial to the receiving aerial. Similar principles apply for receiving aerials as for transmitters as above.

The Half-Wave Dipole

There is only one part of a receiving aerial that is active, i.e. does the receiving and is connected to the TV/radio set. This active element is called the dipole. The simplest design of antenna would consist of a dipole only:

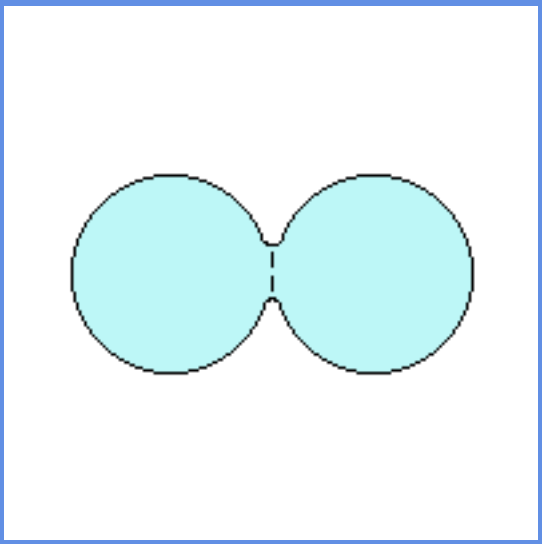


A half-wave dipole

In the diagram above, there are two wires marked 'to receiver.' For UHF and VHF, one wire will be the copper-core and the other the copper braiding of a co-axial cable.

Before we proceed, a quick word about gain. Although having a technical definition, for us 'gain' can mean "the effectiveness with which a receiving aerial receives a signal."

The diagram below shows the reception pattern of a half-wave dipole. The blue area is where the gain is higher than a certain value; the dipole is in the centre:



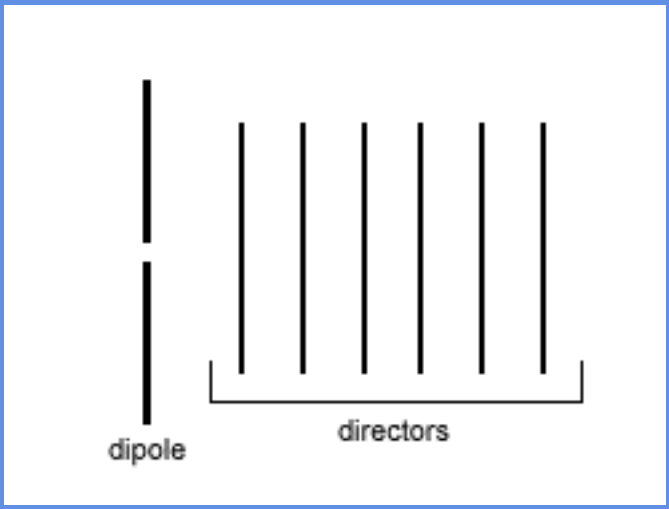
We can change the directivity of the aerial by adding other elements. Any other elements that we add to the basic half-wave dipole are called passive elements and are not connected electrically to the dipole.

There are two types of passive elements:

Directors

Directors alter the directivity of the aerial so that the aerial's gain is improved in front of the dipole. Most aerials have

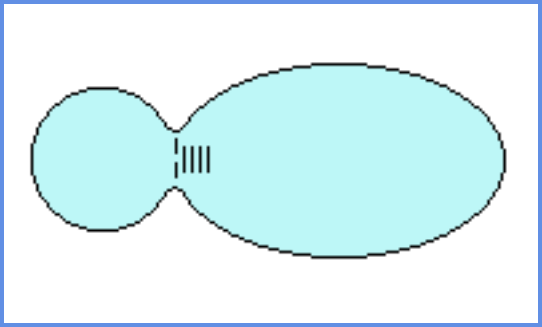
more than one director, and the more directors the aerial has the better the aerial is at picking out the signal from the required source and rejecting signals from other angles.



These diagrams do not show the cross-bar that holds all the elements in place as it does not much affect the characteristics of the aerial.

The spacing between the directors, diameter of the tubing used and the spacing between the first director and the dipole are important in practice but will be disregarded here. The length of the directors governs the bandwidth of the aerial (over which channels it is effective), but suffice it to say that it is about 75% the length of the dipole.

The gain of the dipole with directors in place looks like this:



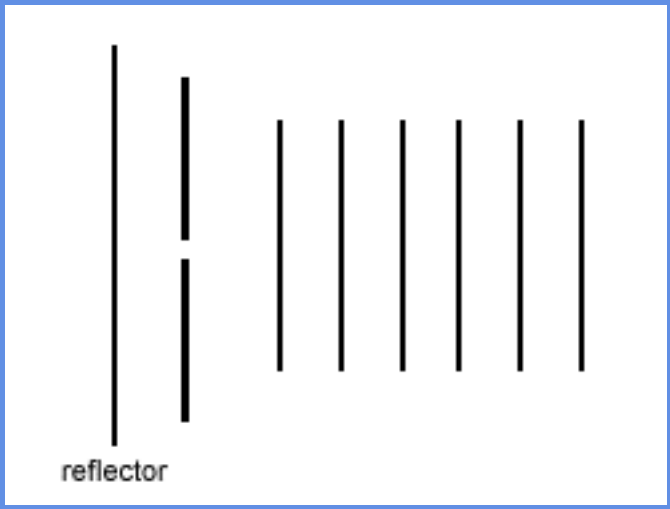
Notice how the gain is now more focused in the direction of the directors.

As stated earlier, the more directors an aerial has the more focused the gain is in the direction of the directors. Every new director added becomes less effective though, and in practice it is only worth adding 18-20 directors to the aerial, as any more than this wouldn't increase the gain very much.

On the diagram above, the aerial still has some gain at the rear - in other words, it can still receive signals from behind. This is known as a low front-to-back ratio.

The Reflector

To improve the front-to-back ratio we can add the second type of passive element, a reflector. The reflector reflects signal coming in from the back of the aerial whilst improving the forward gain.

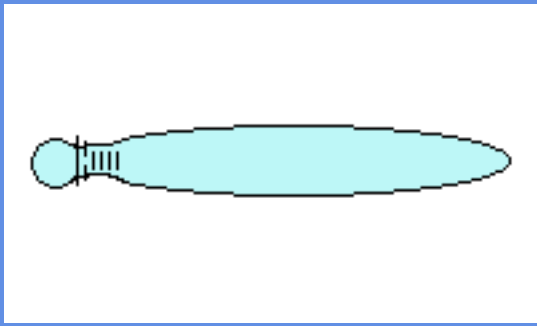


This design is called a Yagi-Uda array, after its creators.

Again, the length, size and position of the reflector affect the aerial's properties, but we won't go into that here.

The reflector can take the shape of a metal plate (with holes in it, making the aerial more impervious to wind) or several rods spaced equidistant from the centre of the dipole.

The result is that there is less gain behind the aerial and more, where we want it to be, in front:

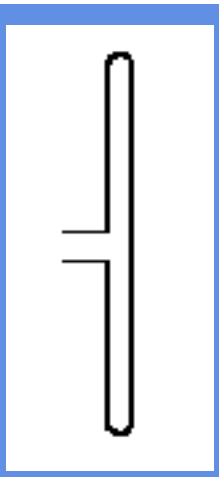


Folded Dipoles

In order to minimise signal loss it is important that the impedance (a sort of resistance for AC) of the dipole matches that of the feeder cable and the receiving set.

The impedance for the type of dipole discussed above is about 75 ohms. More often than not though the impedance needs to be altered to match the cable and receiving set characteristics.

This change of impedance is acheived by folding a rod over so that its folded length is still half-a-wavelength:



Now we know what each constituent part of an aerial is called and what its function is, let's look at some examples in the field.

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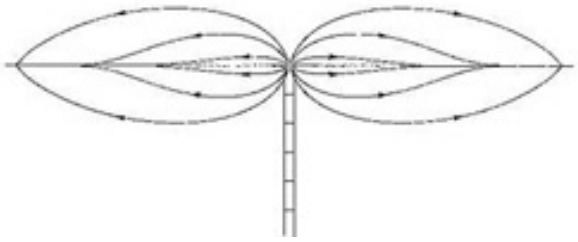
Why an Antenna Radiates

You don't have to know how an antenna works to use one, but getting a handle on this subject can deepen your understanding of radio. Here's a searching look at the mysterious process by which our antennas hurl energy from Here to There.



By Kenneth Macleish, W7TX
QST November 1992

If antennas didn't radiate, we might be well advised to dispose of our ham gear and take up another hobby. Fortunately, they *do* radiate. Our purpose here is to explore the basic principles underlying this wonderful and little-understood phenomenon.

Before starting, let's try a simple test to see how much you already know about antennas. Don't peek at the answers[1] until you have answered all the questions!

1. In a center-fed, half-wave dipole, electrons surge back and forth from one side of the antenna to the other.

True ___ False ___

2. It is possible for a perfect insulator to radiate.

True ___ False ___

3. Unlike ohmic resistance, "radiation resistance" has significance only at the feed point of an antenna or antenna system.

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4. The ground around a transmitting antenna radiates.

True ___ False ___

If you got all the answers right, you probably know more than the writer, and you have permission to skip the rest of this article.

An Imaginary Experiment

Take a pithball about the size of a marble between your thumb and forefinger. If you don't know what a pithball is, or can't find one, a hunk of plastic foam or a ping pong ball will be fine. For lack of a better name we will refer to any of these articles as a pithball. Rub the pithball on your carpet to give it a good electrical charge. Now wave the pithball back and forth in the air over a six-inch distance as fast as you can. The pithball is sending out electromagnetic waves! Let's say you are achieving a rate of 10 cycles of motion per second. If you have placed in the corner of the room a sufficiently sensitive megameter-wave receiver, it will detect a signal when tuned to a frequency of 10 hertz, or to a wavelength of 30 million meters. If you could vibrate your hand fast enough you might even be able to carry on radio communication in this fashion!

We will continue this experiment shortly. Meanwhile let's take a look at what goes on inside an antenna.

Put Your Antenna Under a Microscope

Now, admittedly the average antenna doesn't contain any pithballs. But it does contain hordes of minute, lightweight, electrically charged particles called *electrons.* Many of these are so-called *free* electrons that have broken loose from their parent copper or aluminum atoms and are able to travel more or less freely through the spaces between atoms, under the influence of any electric fields that may be present. The free electrons behave in many ways like tiny pithbalIs.

We know that an electric current in a conductor is simply a mass migration of free electrons. If the current is alternating, as in an antenna, the free electrons in a given locality move back and forth in unison. Evidently, then, any individual electron moves to and fro around an average position, like the pithball in our experiment. Let's see how far and how fast this electron might travel.

Consider an antenna made of #12 copper wire and operating at 14.1 MHz. Each free electron near the surface of the wire is executing 14.1 million cycles of motion per second. Knowing the number of free electrons per cubic inch of copper, the electric charge on each, and the depth of RF penetration into the wire (the *skin depth),* we can calculate the peak speed of an electron at a place where the RMS antenna current is, say, one ampere. The result comes out to be less than half an inch per second. At that rate the electron doesn't move very far during each half cycle of vibration: its peak-to-peak travel is no more than a hundred-millionth of an inch. In the eyes of an electron this distance is quite respectable, being tens of thousands of times its own diameter. The answer to question 1 above, though, is clearly "False." Not one electron makes it through the feed system from one side of the antenna to the other.

We can compute the electron's deceleration and acceleration, which are greatest when the electron is coming to a stop and then starting up in the other direction. At an antenna current of one ampere, these quantities reach more than 50,000 gs! And an accelerating or decelerating charged body, be it an electron or a pithball, is a source of

electromagnetic radiation.

A pithball is a pretty good insulator, and so, presumably, is an electron. So the answer to question 2 is "True": a perfect insulator *can* radiate.

Let's return to our imaginary experiment. This time, instead of waving a pithball at 10 hertz, we'll pretend to vibrate a single electron at radio frequency[2] and examine the resulting fields.

Fields of a Vibrating Electron

The detailed structure of an electron is unknown, but for our purposes it doesn't matter; we can assume that our electron is a little round ball with an electrical charge distributed uniformly over its surface. We determine the fields by solving Maxwell's equations[3] at all distances from the electron, right down to its surface. This analysis is not an exercise for the faint-hearted. We'll skip the details and concentrate here on the results.

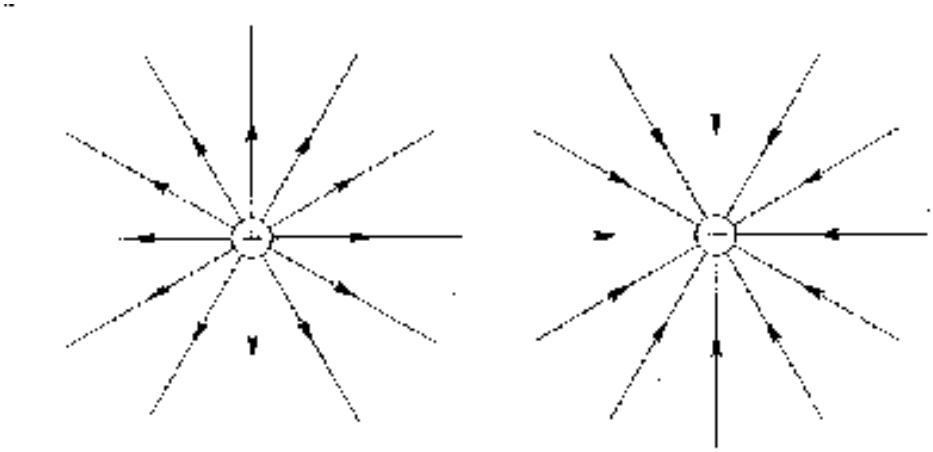

Fig 1 -- Always present around a charged particle, the coulomb field plays a vital role in energy radiation.

The Coulomb Field

Grasping the electron in, say, a tiny pair of tweezers, let's start by holding it still. After a while the only field present will be a stationary electric field that points outward[4] in all directions from the electron. The field lines take the form illustrated in Fig 1 for both positive and negative charges. This is called the *coulomb field.* It is always present, regardless of whether the electron is in motion. We'll find later that the coulomb field plays a vital role in the operation of antennas.

With vibratory motion, two new fields make their appearance.

The Magnetic Field

A moving electron constitutes a current, and a current is always surrounded by a *magnetic field.* As if hitchhiking, point the thumb of your right hand in the direction of the electron's motion; then your curled fingers represent the circular lines of the magnetic field around the electron. Point your thumb in the opposite direction and you see that the

magnetic field reverses, so a vibrating electron gives rise to an alternating magnetic field. At the electron's surface, the magnetic field is almost exactly in phase with the electron's speed, but as we move away, the phase of the magnetic field begins to lag. Out to a radius of 1/6 wavelength, the phase lag is small. Beyond this radius, the lag starts to increase at a more rapid rate and soon settles at 360 degrees per wavelength of distance.

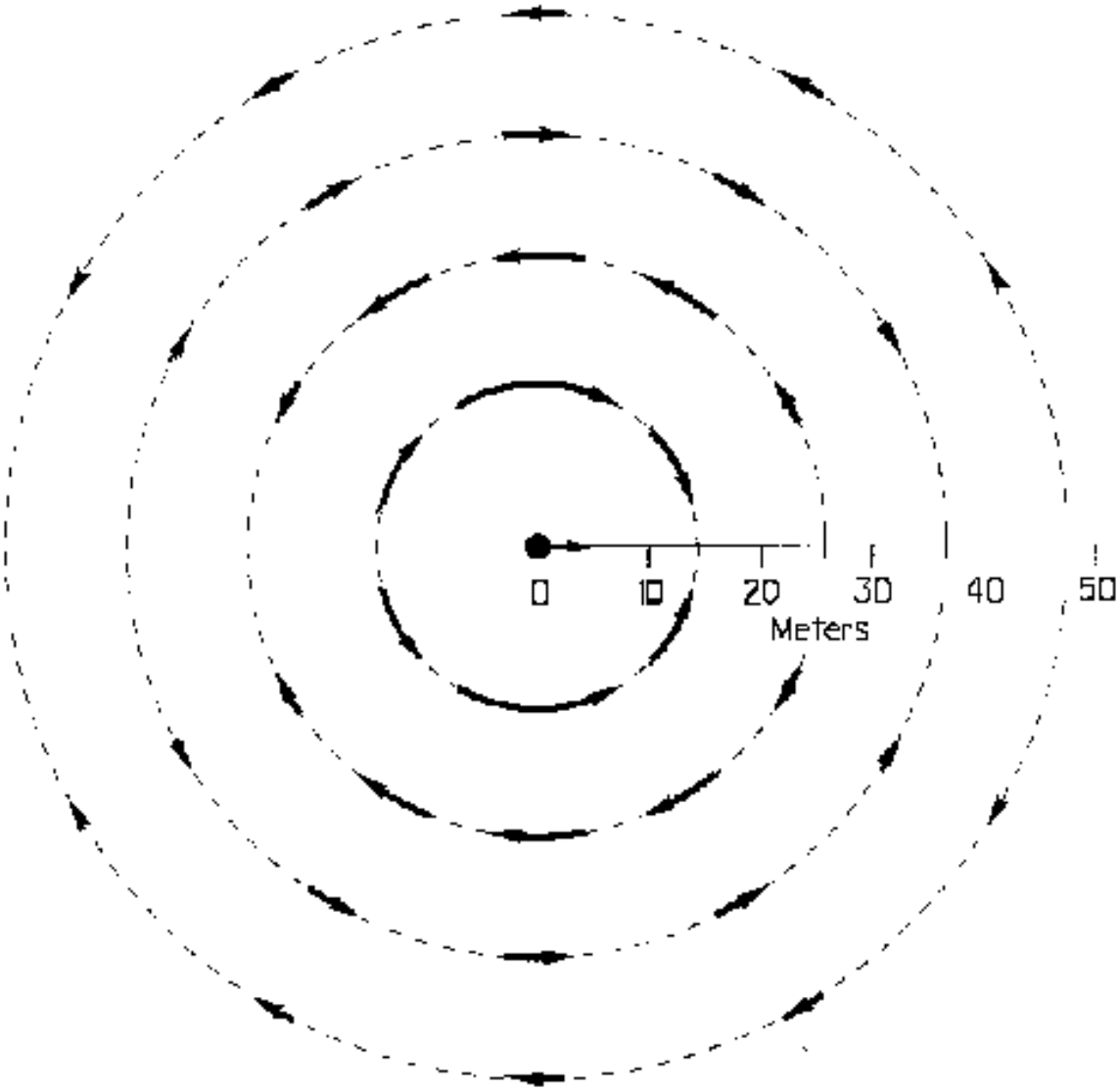

Fig 2 -- The radiation field of an electron vibrating as 14.1 MHz. Each broken line represents a wave crest.

The Dynamic Electric Field

The second new field is an electric field that results from the electron's acceleration. Because of its dynamic origin -- in contrast to the electrostatic nature of the coulomb field -- we'll call this field the *dynamic electric field.* It's useful to regard the dynamic electric field as the sum of two separate fields, one of which is in phase with the magnetic field and the other 90 degrees out of phase. We will call the in-phase component the *radiation field* and the out-of-phase component the *induction field.* It is the radiation field that carries energy from an antenna into the surrounding universe.

Fig 2 shows an area the size of a football field in which a lone electron, greatly magnified, is being vibrated at 14.1 MHz. The figure is a snapshot of the electron's radiation field taken at an instant when the electron is at the center of its travel and moving to the right, as indicated by an arrow. The curved arrows show the direction and strength of the radiation field. The dashed circles represent spherical wavecrests on which the field is at a local maximum. As we go outward from a wavecrest, the field decreases to zero, then reverses and rises again to the next wavecrest. Like ripples in a cosmic pond, these spherical waves are expanding outward at the speed of fight, 300 million meters per second.

At any one point, the radiation and induction fields vary as sine-wave functions of time. About 1/6 wavelength away from the electron (actually 1/[2p] wavelength), in a direction at right angles to the line of motion, the two fields are equal in amplitude. As we move farther away, the induction field falls off so much more rapidly than the radiation field that we soon have essentially nothing but the radiation field. Inside a vacuum-dielectric coaxial transmission line carrying power in one direction, the ratio of the electric to the magnetic field is equal to 377 ohms. In our pure radiation field, which carries power in the same manner even though there are no physical conductors present, this ratio is likewise equal to 377 ohms-a value that is sometimes called the characteristic impedance of space.

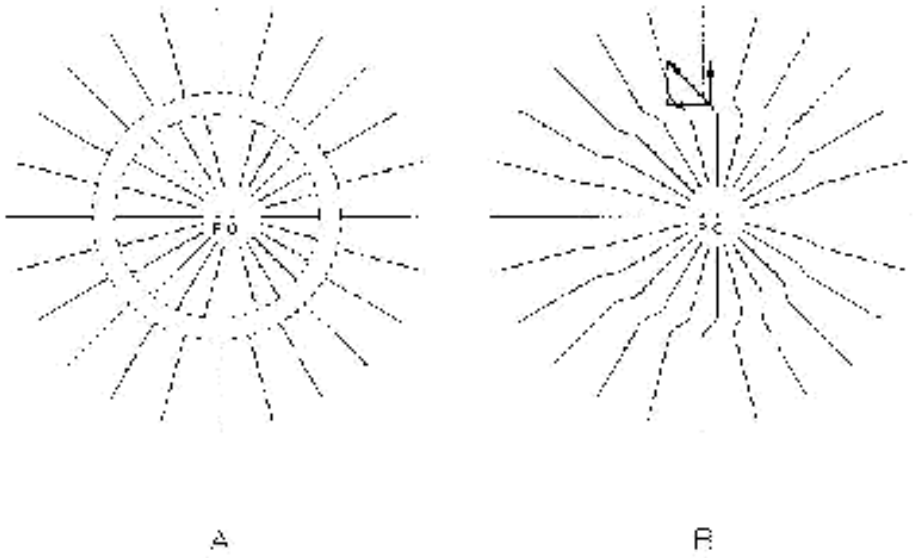

Fig 3 -- The electric field of an electron that was accelerated recently. The resulting disturbance in the field is traveling outward at the speed of light. The transverse component of the disturbed field is the radiation field.

Why an Accelerating Electron Radiates

We've described the radiation field of a vibrating electron, but we haven't yet explained why it happens. The answer is hidden in Fig 3A.

Suppose that, until a short while ago, an electron was held at rest at point P in Fig 3A. It was then accelerated briefly to the right by our tweezers and afterward was kept moving to the right at constant speed. At the present time (which we'll call time zero), the electron is passing point Q.

Fig 3A contains two circles. The larger circle (the outermost broken fine) is centered at P and has radius equal to the distance light would travel in the interval from the beginning of acceleration until time zero. The smaller circle is

centered at the spot occupied by the electron at the end of acceleration; its radius is equal to the distance light would travel between the *end* of acceleration and time zero. As time marches on, the circles evidently grow at the speed of light. The space between the circles is equal to the distance light would travel during the period of acceleration. If the electron moves slowly in comparison with light, as it does in an antenna, the distance it covers during acceleration is small compared to the size of the circles, so the circles are nearly concentric. For clarity we have greatly exaggerated the distance PQ; it too would be very small if drawn to scale. Now we can determine what the electric field must look like at time zero.

Outside the larger circle, the field at time zero is a stationary coulomb field centered on P, as if the electron had never started to move.[5] Inside the smaller circle, the field is a moving coulomb field centered on the electron's present position, point Q. Between the circles the field is intermediate between the fields in the other two regions.

Now connect the field lines across the space between the circles and erase the circles, making Fig 3B. You can see that the electron, while accelerating, gave birth to an expanding electromagnetic disturbance. In the disturbed region, as shown by the arrows, there is a transverse field component -- the radiation field -- in addition to the outward-pointing coulomb field.

The radiation field resulting from a *vibrating* electron, Fig 2, is simply a continuous series of such disturbances caused by successive intervals of changing acceleration and deceleration.

The Bootstrap Forces

The radiation and induction fields of a vibrating electron exist right down to the electron's surface. Since the electron's surface carries an electric charge, and since an electric charge is pulled by an electric field, it's fair to ask whether these fields are able to exert forces on the very electron that is producing them. In other words, can an electron "feel" its own dynamic electric field? The answer is yes. The electron is pulled by its own bootstraps! The tweezers that are providing the motive power must overcome the bootstrap forces.[6]

The bootstrap forces are responsible for two very important properties of a conductor: radiation resistance and inductance.

Radiation Resistance Versus Ohmic Resistance

By our definition, an alternating radiation field is in phase with the accompanying magnetic field. At the surface of a vibrating electron the magnetic field is essentially in phase with the electron's speed, so here the radiation field, and the bootstrap force exerted by it, are likewise in phase with this speed. The direction of the force is such as to resist the electron's motion. It is evident that the force feels to our tweezers like a drag proportional to speed, as if the electron were moving through a viscous fluid. This drag force is the cause of radiation resistance.

An electron moving in a conductor also feels a drag force that is due to frequent progress-impeding collisions between the electron and the atoms in its path. This drag is the cause of ohmic resistance, the familiar R in Ohm's Law.

Both kinds of resistance dissipate energy at a rate equal to the resistance times the square of the current. Of course, energy dissipated this way doesn't actually disappear. An alternating current, flowing against *radiation* resistance, turns electrical energy into radiant energy, which wings its way off into space. Current flowing against *ohmic* resistance transforms electrical energy into heat, which is mechanical vibration of the atoms of the conductor -- the atoms vibrate when they're hit by the moving free electrons.

Radiation resistance varies along the length of an antenna wire, but it is independent of the diameter and material of the conductor. The middle third of a half-wave, 14.1-MHz dipole has a radiation resistance of 3.7 ohms per foot. That's nearly 80 times the ohmic resistance of clean #12 copper wire at this frequency. Closer to the ends of the antenna, the radiation resistance is even higher.

Inductance

At the surface of a vibrating electron, the induction field, being 90 degrees out of phase with the magnetic field, is 90 degrees out of phase with the electron's speed (ie, the current). The bootstrap force of the induction field therefore opposes *the rate of change of current* rather than the current itself. Here we see the underlying cause of the property of inductance. In reacting to this bootstrap force our tweezers deliver energy *to* the electron during acceleration and receive back an equal amount of energy *from* the electron during deceleration. The delivered energy is stored in the magnetic field around the moving electron and is returned when the magnetic field collapses as the electron slows down.

Because the bootstrap force of the induction field is proportional to acceleration, it feels to the tweezers just like mechanical inertia. In consequence the electron has an *effective inertial mass* that greatly exceeds its gravitational mass.

Now let's step back and look at the fields in and around an entire antenna. You will recognize that the basic principles involved apply to any kind of antenna. Because of its simplicity, we will use an isolated, center-fed, half-wave dipole as an example.

The Big Picture

If our antenna doesn't contain any pithballs, it doesn't contain any tweezers either. What is it, then, that causes the free electrons to vibrate? In real life it takes an electric field to move an electron. Inside an isolated straight dipole, the motive power comes from the combined coulomb fields of all the charged particles, positive and negative, in the antenna. We'll refer to this combined field as the *antenna's* coulomb field.

In addition to the coulomb field, the antenna as a whole exhibits a magnetic field that is the sum of the magnetic fields of all the moving free electrons. It also sports a dynamic electric field that is the vector sum of the dynamic electric fields of all the free electrons. As we did with an individual electron, we can separate the dynamic electric field of the antenna at any point in space into two components, one in phase with the total magnetic field and the other 90 degrees out of phase. We will call the in-phase component the radiation field of the antenna and the out-of-phase component the induction field. Right at the antenna, both fields are parallel to the metal surface.

It happens that the coulomb field and the induction field fall off much more rapidly than the radiation field with increasing distance from the antenna. At distances greater than a few wavelengths from the antenna, in what is called the antenna's *far field*, the electric field is essentially pure radiation.[7] Closer to the antenna, we have the *near field*, which is a mixture of the radiation, induction and coulomb fields.

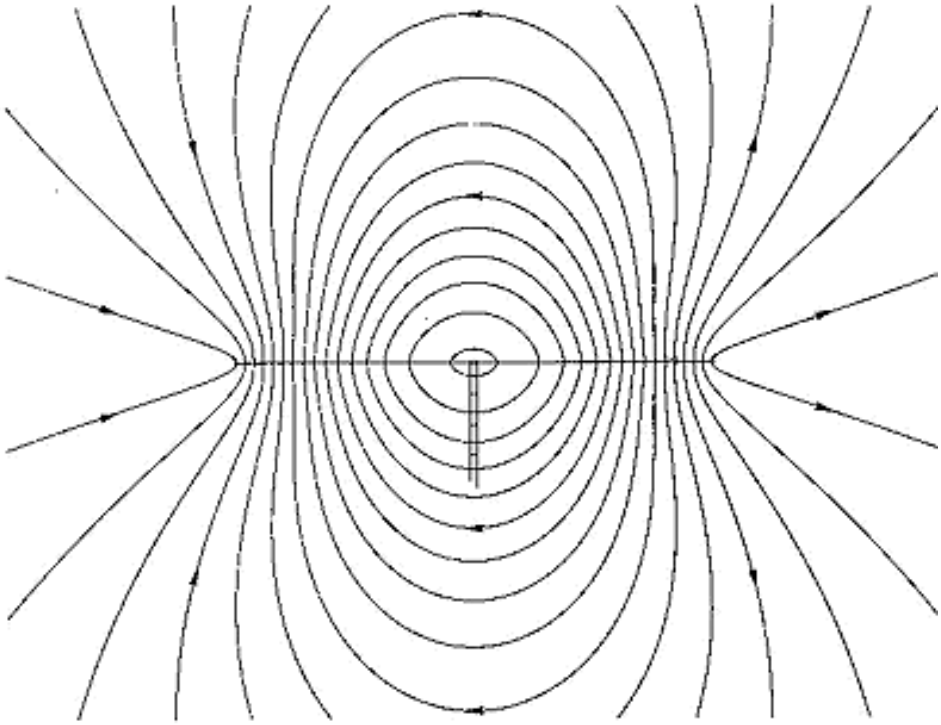

Fig 4 -- The coulomb field at an instant in time around a half-wave resonant dipole A half-cycle later, the polarity, and all the arrows, will be reversed. The spacing between the field lines indicates field strength.

Action of the Coulomb Field

We see in Fig 4 a snapshot of the coulomb field near the antenna. This picture shows an instant in time when the right-hand half of the antenna is positively charged and the left-hand half is negatively charged, as a result of a process that we'll examine in a moment. A half-cycle later, the polarity, and all the arrows, will be reversed. The spacing between the field lines indicates field strength. At the antenna wire or tubing, the field lines are nearly, but not exactly, perpendicular to the metal surface.

On alternating halves of the antenna, the perpendicular component of the coulomb field tries to pull electrons out of the surface. This effort is generally unsuccessful in amateur antennas because the "work function" for copper or aluminum -- the energy it takes to dislodge an electron from the surface -- is too great. If the transmitter power is very high, though, the field may be strong enough to pull electrons out into the air. The result is an exciting (but power-wasting) luminous display called *corona.* But because the coulomb field leans slightly away from the perpendicular at the antenna's surface, it can always pull free electrons *along* the surface.

At one point in the RF cycle, free electrons throughout the antenna are moving to the right at or near their maximum speed. The right-hand half of the antenna thereupon begins to accumulate an excess of electrons, even if no single electron will shift to the right by more than a hundred-millionth of an inch. In the left-hand half of the antenna the departure of free electrons leaves an equal excess of oppositely charged metal ions, which are stationary atoms that have lost an electron. The coulomb field produced by this increasing imbalance of charges now opposes the electrons' rightward motion. By virtue of their effective mechanical inertia (also known as the bootstrap force of their induction field), the electrons coast for a while against the rising force of the coulomb field, which eventually brings them to a

stop and then propels them back toward the left. After the electrons again reach maximum speed, now in the opposite direction, the foregoing scenario is repeated with left and right interchanged. The end result is the vibratory motion of free electrons that causes them to heat the metal and generate electromagnetic waves.

Newton's second law of motion tells us the relationship between the acceleration of an electron and the sum of the forces acting on it. In this case, one of the forces is the pull of the coulomb field parallel to the metal surface. The other two forces are the bootstrap force of the dynamic electric field and the drag of ohmic resistance. According to Sir Isaac, the sum of all three forces is equal to the gravitational mass of the electron times its acceleration. We will assume that an electron is so nearly weightless that its gravitational mass can be set to zero in this equation. Then the three forces must always add up to zero.

Turning this statement around another way, we can say that the dynamic electric field and the parallel component of the coulomb field partly cancel each other, the remaining field being just enough to overcome the drag of ohmic resistance. If the ohmic resistance is small enough to be ignored, the coulomb field is precisely equal and opposite to the dynamic electric field everywhere on the surface of an antenna.

This result leads to a procedure, which amounts to the construction and inversion of a matrix, for computing the current distribution on an antenna. Then, using the principles we've been discussing, we can compute the behavior of the system in detail.

Power Flow Through Space

Electrical engineers use the term *real power* for power that flows in one direction past a given point. They also speak of *reactive power*, which is power that flows back and forth in alternating directions with a net flow of zero in any one cycle. The radiation field of an antenna transmits only real power, which travels out toward distant localities without ever reversing direction. The induction field carries only reactive power, and the coulomb field carries both real and reactive power. Again we'll illustrate this with an isolated, center-fed, half-wave dipole.

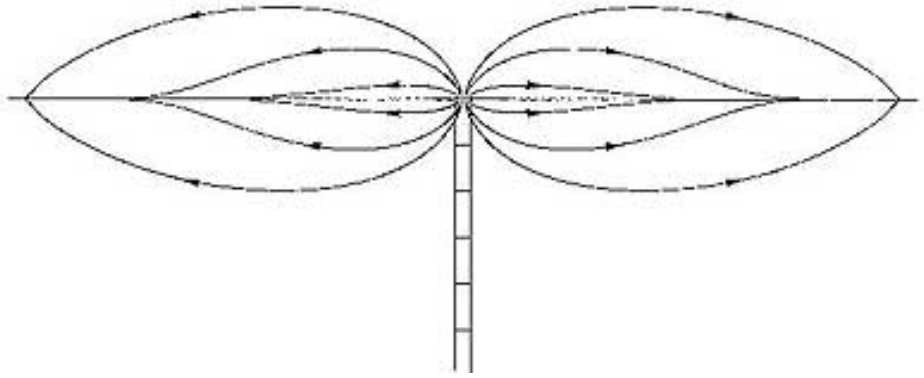

Fig 5 -- Flow of power from the feed point of a half-wave resonant dipole. The coulomb field around the antenna conductors carries power *through the space surrounding the conductors.*

Real Power Flow

How do you suppose your hard-earned RF power gets from the feed point to the rest of the antenna? You might think that it flows out through the wire or tubing of the dipole, but actually this real power is carried *through the*

surrounding space by the coulomb field. Some lines of power flow are plotted in Fig 5.

As you see in the figure, when power coming up the feed line reaches the antenna it spews out into the air in all directions. It then arches to the right and left on curved paths that eventually intersect the antenna. At the points of intersection the coulomb field donates its real power to the free electrons, making up for the energy they are losing to ohmic resistance and radiation resistance.

Reactive Power Flow

During one interval in the RF cycle, the charge on the antenna reaches a maximum, and the current and the magnetic field go through zero. A quarter of a cycle later, the reverse is true. In the first interval, the antenna is surrounded by a cloud of electrostatic energy stored in the coulomb field. In the second interval, the coulomb field has disappeared, and we find the same energy stored in the magnetic field. The energy stored in the coulomb field is used in accelerating the effective inertial mass of the free electrons, which by their motion create the rising magnetic field. Energy thus moves from the coulomb field, via the induction field, to the magnetic field, only to move back again during the next quarter cycle as the magnetic field collapses. This is reactive power flow, with a net of zero.

You can think of the cloud of electrostatic energy as energy stored in the distributed capacitance between the two halves of the antenna. Similarly, the stored magnetic field energy can be thought of as energy residing in the distributed inductance of the antenna wire. If power is suddenly cut off by short-circuiting the feed point, the antenna doesn't stop radiating right away. Instead, it oscillates with diminishing vigor at its resonant frequency until the energy stored in the fields has been dissipated in ohmic resistance and radiation resistance. Our antenna can be accurately described as a resonant circuit made up of distributed capacitance, distributed inductance, and two kinds of distributed resistance.

Antennas and Non-Antennas

What do the following items have in common: a dipole antenna, a radar dish, the ground around a transmitting antenna, a coil and capacitor in parallel, and a wire carrying music to a loudspeaker? One answer: to a greater or lesser degree, they all radiate!

This is true because in operation they all carry time-varying currents and, consequently, accelerating electrons. The dipole antenna is an example of a distributed circuit that owes its existence to the fact that it radiates well. It is designed for efficient conversion of electrical energy into radio waves. But any system of conductors that carries varying currents behaves in accordance with the principles described earlier. The same processes, including radiation, take place whether we call the system an antenna or something else. For example, what we generally refer to as reflection from a conducting surface (the radar dish, the ground) is actually radiation from free electrons set in motion by incident electric fields.

In Sum

Perhaps the foregoing intuitive introduction to classical electromagnetic theory will inspire further development and refinement of the ideas presented. In any event, I hope you'll be able to contemplate an antenna, or a non-antenna, with a warm feeling for all the interesting things going on in and around it!

Notes

[1]1--False. 2--True. 3--False. 4--True.

[2]The classical electromagnetic theory we use in this article is valid at frequencies from dc through microwaves. At still higher frequencies one must venture into the realm of quantum electrodynamics.

[3]For some pertinent theoretical background see Feynman, Leighton and Sands, *The Feynman Lectures on Physics*, Vol 11 (Addison-Wesley Publishing Co, 1964).

[4]To get an outward-pointing field we must take the charge on the electron to be positive instead of negative as is actually the case. I find it easier to visualize the interaction of electric fields and electrons by thinking of the electron charge as positive. In the end it doesn't make the slightest difference. After all, Ben Franklin, on rubbing an ebony rod with cat's fur, could just as well have defined the charge induced on the cat to be positive. Life might be simpler today if he had.

[5]If you were observing from anywhere outside the larger circle at time zero, the electron would appear to be at point P instead of its actual position, point Q. That's because, according to Einstein's theory of relativity, the news that the electron has started to move can't propagate faster than the speed of light. If you were anywhere inside the smaller circle, the electron would appear at time zero to be in its actual position.

[6]The electron is pulled by its own coulomb field too, but equally in all directions, so the net effect is zero.

[7]In summing the radiation fields at a given point in the far-field region from all the free electrons, we must take into account the differing phases of the source electrons and the varying propagation times. The resulting phase variations cause partial reinforcement or cancellation of the fields, to a degree that depends on direction from the antenna. This effect is the basis of the very extensive subject of antenna radiation patterns, which we will not discuss further here.

Ken Macleish received his first Amateur Radio license 60 years ago at age 12. He now has an Extra Class ticket and still enjoys ham radio, especially CW ragchewing.

He holds a BS in physics (1939) from Caltech and a PhD in physics (1943) from the University of California at Berkeley. He was with Tennessee Eastman Company at Oak Ridge, Tennessee, during World War II, then transferred to Eastman Kodak in Rochester, New York, where his most recent position was assistant director of research and engineering in the Kodak Apparatus Division. In 1962 he joined the Perkin-Elmer Corporation in Norwalk, Connecticut, as a vice president of engineering. He retired in 1970 to enjoy the climate and scenery of Arizona.

Ken has authored QST articles on frequency counters and frequency measurement. Also, over the years, he's had a nagging desire to understand why a radio antenna works. Combining this desire with a background in physics, a hobby of ham radio and an interest in computers, Ken has been working off and on for some time to help dispel some of the mystery surrounding this subject.

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Construction Practices

KQ6RH

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Ray Jurgens

(Up-Dated 2/25/2000)

Introduction

The construction of light weight structures that support wire antennas requires some definite insight into the behavior of the materials. I shall consider only constructions using fiberglass tubes that telescope and the guy lines that are required to stiffen the structures. In some cases, the guy lines are actually the antenna wires, to the mechanical properties of the wires is also important. To understand the structural properties of fiberglass tubing, you should read Properties of Fiberglass Rods and Tubes. This section should give you a first hand understanding of the mechanical properties of this material and a good feel for the amount of bending that can be expected for various diameters and lengths of tubing. The total bending of composite telescoping spreaders made from the standard sizes of tubing can be estimated by considering the loading of the outer most sections at the beginning of the next larger section and working inward so long as the total angle of bending remains small. Small for practical purposes is about 20 degrees. In the case of most light structures, the bending will be more than is desirable, and guy lines will be required to take some the load thus reducing the total bending moment along the beam or spreader. In the end, the maximum loading is controlled by the tendency of the spreader to buckle due to the total force of the guys along the axis of the tubing or rods. This limiting force is easily computed given the geometry and the modulus of elasticity of the material. The remainder of this section considers mostly construction practices that have been found to work well and save time in assembly and disassembly of the antennas. As most of the antennas considered in this web site can be considered as portable, assembly time is often an important factor.

Telescoping Spreaders

The standard tubing sizes available from Max Gain Systems begin at 1/4" OD and work

upward to about 2" in 1/4" increments. The walls are slightly less than 1/8" thick allowing the sections to telescope. In the case of the 3/4" tubing, the OD is slightly too large, and it is not possible to insert this size into the 1" material more than about a foot without getting it seriously stuck. This works fine as a fixed length spreader, but you can not vary the length. I shall focus on the construction of light weight antennas using only the 1/8", 1/4" and 1/2" sizes for the spreaders and the 3/4" and 1" sizes for boom and upper mast sections. These materials are sufficiently light that all antennas I have considered here can be easily constructed and erected by a single person.

The standard procedure used to assemble telescoping aluminum sections works equally well here. Six inches to one foot of overlap is always desired between sections. The sections are locked in place by using standard hose clamps to compress the tubing at the tip. For this to work, a slit must be cut at the tip of each section to be compressed, and the depth of the cut should be about twice the diameter of the tubing. So, for 1/2" material the cut should be about 1" and the hose clamp mounted exactly flush at the tip. The cut can be made most easily using a hack saw. you should be careful not to breath the dust from this operation, i.e., I suggest you use one of the paper filter masks while doing the sawing. You may find it difficult to find hose clamps smaller than 1/2" in most hardware stores, however, the 1/2" size will normally compress down to 1/4" but that is about the limit. The 1/8" solid rod can be used for micro extensions to the spreaders, but these are normally limited to about 6" in length. As a result, the maximum length of the spreader is limited to about 16' when using 8' sections of 1/4" and 1/2" materials. So the largest practical structure might support a full-size rotatable 20 meter dipole. A more useful limit for quads and planar wire beams is about 12'.

Supporting Spreaders

After considering many standard procedures for supporting spreaders, I decided that quick assembly was more important than cheap. There are several inexpensive ways to mount spreaders, for example, you can use a 1 or 2' section of 3/4" fiberglass rod in the center that can be U-bolted to a plate or hub. The plates and hubs are normally made of aluminum or hardwood. I experimented with both materials and found that neither suited me for this application. After discussing the problem with a good machinist, he suggested that we try PVC. We initially constructed four hubs from this material for testing, and found that it is sufficiently strong, light in weight, dark in color, easy to machine, and holds tolerances well. Currently, we are making these available in one size only, that is, for 1/2" spreaders with a 1" boom or mast. Stainless steel set screws are used to lock the spreaders into the sockets and to lock the hub to the most or boom. The initial hubs have been in use for two nearly years at this point and show no degradation of the material other than a slight change in color. See our Products Page for details and cost. These hubs permit rapid assembly and disassembly using a small Allen wrench. We have paid considerable attention to the accuracy of the machining. The spreaders

are held accurately at ninety degrees and the length from the center of the hub to the tip is exactly one inch longer than the spreader when the spreaders are fully inserted in the sockets. These tolerances are important to maintain so that the lengths of the guy lines can be accurately calculated when they are needed.

Guy Lines

We have experimented with and tested guy lines made of three materials at this point. The decision of which to use depends upon a number of considerations such as visual impact, the amount of stretch that can be tolerated, the degradation over time related to weather and ultraviolet, and the expense of the material. The following materials have been tested:

Material	Diameter	**#Test**	**E, Modulus of Elasticity**
Nylon Fishing Line	25/1000"	40	
Black Dacron Rope, DR33225	3/32", ~ 105/1000"	>600	
Bonded Kevlar, Size 207	25/1000"	>	

Table 1

Properties of Guy Line Materials

Of the three materials, the Black Dacron is most visible due to its greater diameter, the Kevlar is next due to its gold color, and the Nylon is least visible of the three if the material is clear or lightly tinted. Of the three materials, the Kevlar stretches the least for a given force and does not creep with time. In general, it is difficult to measure the tensile strength of these materials, as it depends upon how the materials are tied or bonded to the supports. These materials will always break at knots, so, that tells you that tying knots in the material is not the best way to form the tie-loops. After considerable testing we found that crimp fittings designed for fishing leaders worked well for the monofilament Nylon lines. These work less well for the multi-filament line, because the line flattens easily and requires the crimp to compress fully which is often doesn't. This results in slippage under tension. In the case of Kevlar, we found that most of the failures could be prevented by double crimping and tying a double half-hitch behind the crimp. The the larger diameter Dacron material can be handled with standard aluminum crimp fittings designed for wire cable. Again, the crimp fittings have to be compressed evenly across the tube. It is a good idea to do a few practice crimps and test them under tension just to be sure you have the technique under control. Once you are sure of this, it is time to make some

guy lines of specific length.

The easiest way to make guy lines of accurate length is to fabricate them on a wood board about 8' feet long using two finishing nails separated by the desired length if the lengths are less than 8'. Longer spans can be looped back one or more times, i.e., the nails are spaced at some sub-multiple of the required length. We have found that stainless steel leader clips are very useful for attaching guys to the eye-bolts. The leader clips permit easy assembly and disassembly-assembly of the structures and provide a simple method of attachment that is stronger than most other methods. The combination of the crimp fittings and the leader clips permits the guys to be assembled under tension. As shown in the figure below. The second figure shows

The computation of the length of guy lines is normally not very difficult in the case that the assembly consisting of a hub and spreaders is planar. However, even in that case, one must make allowance for the distance of the attachment points form the spreaders, guy posts, and any other attachment structure that changes the distance between the two points of attachment. If you use snap clips, you must include them in the length of the guy. As an example, consider calculating the length of a guy line that spans the distance from a central guy post to a point near the tip of an 8' spreader. The lines are to be secured with snap clips to eye-bolts that extend perpendicular to the center post and to the spreader. Suppose the guys are attached to the center post at a height of 36" above the plane of the antenna. The eye-bolt extends 1.5" from the center of the post. The attachment to the spreader is set at 93 inches form the center of the hub, and the eye-bolts extend 1.25" above the center of the spreader. The spreaders are to be planar (not bent). The offsets due to the eye-bolts define a slightly smaller right triangle whose length, L, and height, H, are as follows:

L = 93 - 1.50 inches

H = 36 - 1.25 inches

So, if S is the required length of the guy line, the required length is:

S = sqrt (L^2 + H^2) = 97.9"

In the case where the spreaders are bent out of the plane, the computation of the guy lengths is a little more complicated. This is because as the spreader is bent by pulling on the guy point near the tip, the length of the guy point moves up and out of the plane and closer to the center guy pole. If you choose the amount to bend upward out of the plane, you then must know how far that point moves toward the center as well as the angle of the bend at the guy point on the spreader. Calculating these quantities is possible but a bit complicated, so it may be simpler

just to bend the spreader and measure the length from the guy point to a point perpendicular to the guy pole or directly to the attachment point on the guy pole. However, the table below gives the approximate length of the tip of the spreader perpendicular to the center of the guy pole, the elevation of the tip above the plane of the hub, and the approximate force required to pull the spreader into this shape when guyed to a 3' guy pole.

All materials used for guy lines stretch, and the amount of tension in the line determines how much it stretches. Different materials of various weights stretch different amounts, so you may want to measure the amount of stretch. Nylon fish line stretches quite a lot while Kevlar thread hardly stretches at all. The stretched materials also grow in length under tension, for example, a Nylon line cut initially to 100" may creep to 104" when to about 3 kg or 6.6 lbs. It will also almost return to its original length when left un-tensioned. Because of this behavior, it is very difficult to measure the modulus of elasticity except by dynamic means. On the other hand, a similar length of Kevlar thread size #207 will stretch only about 3/8" when tensioned to 3kg and returns to the same length when the tension is removed. However, even the Kevlar deforms slightly over a period of weeks by about 3/8" to ½" for a 100" length. This means that you should provide for some room for adjustment when setting the locations of the guy points. You should assume that line lengths will grow a bit and provide space to extend the guy points. It is also possible to alleviate the stretching by pre-tensioning the material before cutting the lengths, and this is especially true for Nylon. In the case of Kevlar, the tension may take several days to weeks to stretch the material. Pre-tensioning also is commonly done with copper antenna wire, as it also grows in length in a similar manner. So, it is a good idea to pre-tension all materials before cutting. The only exception to this rule might be the Kevlar.

Fortunately, it is easy to compute the amount of stretch for a given material with simple tools. If you have a fishing scale, a pulley, a long board and a few nails, you have every thing you need. Or you can use the same materials that we use for which the numbers are known. (to be completed)

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W2IK Presents:

"OVERLOOKED ANTENNA CONSTRUCTION TIPS"

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By Bob Hejl - W2IK c2004

Many ham radio operators enjoy the triumph of constructing or repairing their own antennas. As guidelines, they often use one or more of the many antenna designs which are available in both books and on the Internet. Sometimes they might even take a basic design and adapt it to increase it's efficiency or to make the antenna fit within the physical limitations of their property. It is unfortunate that in these designs little thought is ever given regarding material selection beyond it's cost, weight and strength. There is another important detail that must be taken into consideration during selection and before construction is to begin. The best antenna design using the strongest materials available may be a fruitless effort if you expect your antenna to perform for an extended period of time. One of the most overlooked antenna construction considerations involves the action that usually occurs when two dissimilar metals are brought into physical contact (mated) during the assembly process and exposed to the weather. There can be a serious onset of a type of corrosion at their respective contact points.

This corrosion, also called "galvanic corrosion", can create havoc in as little as three months. Galvanic corrosion is a process that occurs when certain metals are connected to each other and moisture, such as rain, condensation or even high humidity, is introduced. When all of this is present, electrons flow between the two metals (also called "ionic flow"), which causes the most chemically active metal to change. When galvanic corrosion occurs, it can usually be identified by characteristic surface blistering or white powder corrosion around dissimilar metal points.

If left on its own, the corrosion process will continue, eventually pitting of the metal will occur and electrical or RF conductivity will fail. It's problems can be magnified when an antenna is used in humid, windy, intense heat or cold (thermo-galvanic corrosion) or other harsh conditions. Even an antenna encased in a fiberglass shell, such as a vhf or uhf repeater antenna, can suffer from the effects of this corrosion.

There is a direct relationship between various types of dissimilar metals when they are mated. Some dissimliar metals, such as copper and brass, when mated cause very little corrosion. There are other metals, however, that react most harshly when matched. Zinc and brass, for example, will cause corrosion with the zinc metal quickly breaking down. If you wish to prevent galvanic corrosion the best means, of course, is to use the same metal throughout your construction. If this is not possible your next course of action would be to assemble materials that have a close relationship on a galvanic metals table. Here is a descending list of metals and their associated relationships in order of the most "noble" or least active.

least active ***gold***

silver

silver solder

bronze

copper

brass

nickel (plating)

tin
lead
lead-tin solder
stainless steel
iron/steel
aluminum alloys
aluminum
zinc-galvanized steel

most active ***zinc***

When choosing materials you should try to pick a metal part made from the type closest to the other metal parts you will be using as charted on the above list. It is very important to consider EVERY part of your antenna including clamps and washers. For example, if brass screws are used to hold aluminum tubing in place they will cause a headache when corrosion takes hold. The better choices would be either stainless-steel screws to secure aluminum tubing or brass screws to hold copper tubing.

Remember that if you are refurbishing an antenna system you might be locked into using a hard to find piece of hardware based upon the original construction materials. Some antenna companies skimp by using the cheapest, but improper mating metals. For example, if you are servicing a tall two-meter base or repeater antenna and you notice that the antenna company used stainless steel couplers and set screws to link two sections of brass or copper rods it would be wise to replace the couplers and the set screws with brass components. If you don't, you might find that the antenna will exhibit intermittent high swr when unseen corrosion takes place between the set screws and the rods. If any wind flexes the antenna the flexing will cause the corrosive joint to change the swr and it will not only effect your signal but will also stress your transmitter as the load it sees will constantly change with the wind. You don't want to "lose" a set of repeater output transistors because of antenna joint corrosion that could have been prevented. It's worth the extra time to locate hardware made of either the same metal or one closest to the antenna metal as outlined on my chart. In doing so you will avoid having to re-fix the problems caused by improper repair.

If you need to clean off a metal surface it is important to use a "like metal" brush to do so. Using steel wool to clean a brass rod that might be part of a vhf/uhf repeater antenna will embed small particles of steel into the rod and create a corrosive atmosphere. Use a small brass brush instead. Be aware that even when you solder, the type of solder used (silver solder or tin-lead solder) needs to be considered based upon what you are bonding. When you construct an antenna also be aware that the surface area of each contact point is very important. Making one small point of contact, using one set-screw for example, will increase the possibility of galvanic corrosion creating problems. This is compounded by the fact that any RF energy will have to pass through a smaller contact point causing additional resistance. If you must assemble two pieces of your antenna either design a larger overlap or use two or MORE flat tipped set-screws to hold the joint. Do not use plated set-screws because as soon as they are tightened they will lose some of their plating at the contact point possibly exposing the metal below the plating. This may cause a corrosion problem unseen to the naked eye. Multiple set-screws within the same collar are a must. This not only reduces the problem but also makes a better physical connection that won't be as easily loosened due to winds or vibrations.

Whenever you are inspecting an antenna for refurbishing, the tiny corrosion on the contact point of a set-screw may go un-noticed. Many problems can arise due to galvanic corrosion including high swr, lowered output power, reduced receive capability, introduction of intermod in vhf and uhf operations, induced static in receive, an increase in "stray current corrosion", "telluric effect corrosion" and harmonic interference being generated. If you find it necessary to combine metals not closely linked on the above list, use an intermediate metal, such as if you must attach a copper wire to aluminum tubing, use stainless steel or nickel washers and a stainless steel screw or a tinned solder lug with a stainless steel screw. This isn't the best solution, but it will reduce the chance of corrosion. The best answer is to always use "like metals" where possible.

In every case of bonding or connection, in order to reduce the galvanic corrosion, coat all of your antenna's joints and connections. This can be done with a joint sealing compound, but it is important that the compound is non-hygroscopic (not waterbased) and does not contain any aggressive ions that can be leached during service. Spray or paint on coatings that contain zinc or aluminum powder should be avoided. Epoxy-based coatings are much better. If possible, use a clear type as this will aid in joint inspections. If, during any inspection, you see a peeling of the sealer, remove it completely and re-apply. These epoxy coatings can be obtained at most marine supply companies as they are made for both marine and atmospheric applications. Make sure that the surface you wish to protect (the joint) is clean and dry. The spray type allows you a quick seal coating "in the field". In humid conditions, use a hair dryer to remove any mositure. You do NOT want any moisture trapped under the coating. If you are working with tubing, remember that moisture can be introduced through the ends of the tubing into the interior of the joint so make sure the tubing ends are also sealed. The less moisture the better the joint will be protected. BEFORE YOU COAT: Make sure that no corrosion has taken place at the joint. If it has, the corrosion that has started will continue. Take the time... inspect, clean and re-assemble any questionable joint. Also be aware that even when bolting your antenna to a tower you need to use the right hardware. Refrain from using stainless steel "U" bolts to connect an antenna to

an aluminum tower even though the two metals are close on the list. Use aluminum hardware or galvanized "U" bolts. So the next time you plan on building or repairing that antenna, consider your material selections accordingly.

BEFORE YOU LEAVE:

While I have your ear there is something else you should also know. This same principle also should be applied when repairing or installing a grounding system for your shack. Don't use copper grounding braid and attach it to steel (or even copper-clad steel) ground rods!! This creates a wonderful corrosion point that not only affects your system from a safety standpoint, but also RF may be reintroduced back into your shack and give you a nasty RF burn. Ouch ! Your best bet: Use several 1/2 inch copper tubing sections to act as ground rods, then attach your copper ground wire or braid using copper or brass clamps or wrap it around the tubing and solder it using silver solder. Another reason for doing this is that the copper tubing has more surface area exposure to the ground and is, of course, a better conductor. Use more than one and stagger them so they are not in a straight line and space them differently and not a fractional wavelength such as 1/4 wavelength from each other.

page created with Easy Designer

The Antenna Elmer

Here you will find information about wire antennas as well as directional beams.

HF / VHF / UHF dipoles

Other wire antennas

Design your own dipoles and inverted vees (Metric)

Design your own ground plane (Metric)

Design your own 5/8 wave vertical antenna

Design your own quad (Metric)

Folded dipoles

Design your own dipoles and inverted vees

Design your own ground plane

Design your own half wave vertical antenna

Design your own quad

2 meter 2 element direct connect beam

HF satellite antenna (10/15 meter beam)

2 meter 3 element direct connect beam

VHF / UHF direct connect beams

Fan-tenna...An antenna made from a fan?

Quarter wave line matching transformer

Calculate electrical length of a line

Decimal to fraction conversion chart

Determine the power output to your antenna

Determine the efficiency of your transmission line

How long should my coax cable be?

Frequency scaling

Convert from feet to meters

Convert from meters to feet

Antenna construction tips

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Formulas To Design Your Own Dipoles And Inverted Vees

Most hams are familar with the center fed, half wave dipoles and inverted vees that are very popular and easy to build. This formula will give you a starting point to make these antennas. There are several factors that affect the resonant frequency of any antenna. Some of these factors are: the height above ground, the diameter of the wire, nearby structures, the affects of other antennas in the area and even the conductivity of the soil.

If you've ever used some of the other antenna design programs you may realize that the formula for these types of antennas vary from about 476/f Mhz to 490/f MHz depending on the band and the height above ground!

Fortunately there is a **standard** formula that can be used as a starting point in your design. For a center fed, wire dipole, this formula is 468 / frequency in megahertz. I've always cut the antenna a few inches longer which would allow me to trim the antenna in order to obtain a 1:1 match.

The inverted vee has always been about 3 - 5% shorter than a dipole at the same frequency. I used 4% as a constant in the calculation. You may change this value to any number that you like to use.

This page uses the standard formula, 468 / f MHz to calculate dipole lengths. You may change this number if you know of a better number to use as your starting point.

Enter the formula for the antenna calculation

Divided by Freq MHz

Percent smaller for the Inverted Vee

Your dipole's total length is feet

Each leg of the dipole is feet

Your Inverted Vee's total length is feet

Each leg of the Inverted Vee is feet

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AMANDX PRESENTS

DIPOLE ANTENNA

INTRODUCTION-The longwire antenna is a very effective antenna for the listener who wants to cover all of the shortwave bands from 530 KHZ to 30 MHZ. However if you have some favourite frequencies that you listen to on a regular basis you may wish to consider a dipole antenna. This antenna is a fairly easy to construct antenna and will give you better reception on the frequency it is cut for. Think of a dipole as a longwire that has a insulator in the middle.

FREQUENCY-A dipole antenna will not only work well on the frequency it is cut for, but also for the multiples of that frequency. For example if you cut a dipole for 7.0 Mhz will also work well on 14 Mhz, 21 Mhz and 28 Mhz. This way if you can pick and choose your frequency you can make one antenna work on two or three bands.

LENGTH- To find out how long the antenna should be all you have to do is fill in a simple formula:

FREQUENCY IN MHZ divide by 468 = LENGTH IN FEET

That is the only formula you need ever know to build a dipole antenna.

Formulas To Design Your Own Dipoles And Inverted Vees

This page uses the standard formula, 468 / f MHz to calculate dipole lengths. You may change this number if you know of a better number to use as your starting point.

Enter the formula for the antenna calculation

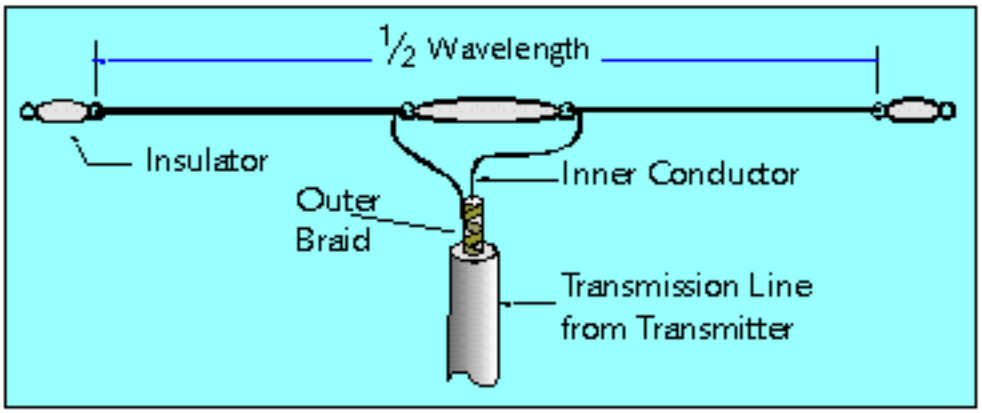
Divided by Freq MHz

Percent smaller for the Inverted Vee

Your dipole's total length is feet

Each leg of the dipole is feet

Your Inverted Vee's total length is feet
Each leg of the Inverted Vee is feet



dipole antenna

CONSTRUCTION- Once you have selected a frequency and calculated the length of wire you need add two feet to this length. This is done so you will have six (6) inches of wire at each end to wrap around the insulators. Once you have this extended length of wire cut it in half. This will give you both sides of the dipole.

Attach an end insulator to one end of each piece of wire. You can use the egg shaped insulators sold by many radio supply stores or make your own out of a piece of plastic. This can be done by cutting a piece of heavy plastic or plexi-glass to a size of about six (6) inches in length and about 2-3 inches wide. Drill a small hole one to one and a half inches from each end of the plastic to wrap the wire around. It is best to solder these connections and wrap them in a sealant tape to keep the effects of the weather from harming them.

The other free ends are attached to a centre connector which you can buy with a built in coaxial cable connecter, or make your own. This will look similar to the end connectors but you will have to find a way to secure the coax lead wire to the insulator. If you build your own when you attach the coax to the ends of the wires, insure that you solder and wrap the connections. One wire will go to the centre of the coax, while the other wire will go to the shielded braid of the coax. This will give you a perfect half wave dipole. You should also wrap the coax fitting of the commercially available centre insulator to keep water and other moisture out. Moisture will ruin the connections on any type of insulator and make the antenna less effective or at worst useless.

MULTI-BAND DIPOLES- As was stated above you can use the dipole on the harmonics or multiples of the frequency it is cut for. However if you are short on space you can build a multi-band dipole. This way you will get an antenna that will operate on several frequencies. Instead of using a single strand of wire you can use wire that has several insulated wires in it. These MUST be insulated wires to insure that they do not touch each other. You then cut the top wire to be the longest, the second wire to be the second longest, the third wire to be the third longest etc.. Only the longest wire is attached to the end insulators and all wires are fed to the centre insulator to attach to the coax feed line.

INSTALLATION- Once you have the antenna cut all you have to do in put it between two masts. Make sure that you use the free side of the end insulators to attach some rope. Tie this rope from the end insulators to the masts. Leave some slack on the antenna. If you pull too tight the antenna will break in the wind or if snow and or ice should coat the antenna. KEEP AWAY FROM OVERHEAD WIRES!! Keep away from these as should the antenna ever come into contact with an overhead wire you will do permanent damage to your radio if not yourself. All you have to do is feed the coax to your radio and listen to the stations come in. It would be best to install a lightening arrester in the coax feed line to help protect your receiver. These are available from many radio supply stores. Follow the instructions carefully! In areas where thunder storms or snow storms are common a lightening arrester is a must for safety.

You can install them flat or at an angle. If at an angle they will be more directional the direction taht they are sloped. You can also istall them as an inverted V shape. This dipole has a higher center with lower ends to save on space in smaller back yards. All three versions work well.
FURTHER INFORMATION- If you want more information on dipole antennas and for that matter all types of shortwave antennas look for these books:

Easy-up Antennas for Radio Listeners and Hams by Edward M. Noll

Joe Carr's Receiving Antenna Handbook by Joe Carr

The Easy Wire Antenna Handbook by Dave Ingram

Practical Wire Antennas by J. Heys

These books are orientated towards the Shortwave Listener more so than the Ham operator. The first two above are probably the best for the beginner and the more advanced, but not technically minded. They put forward a lot of information in a manner everyone can understand. Ed Noll's Easy-up Antennas for Radio Listeners and Hams even comes with tables giving you pre-calculated lengths for many types of antennas. There is lots of tips on making antenna construction simple but effective. It does not come any easier than this.

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Remember On A Clear Day You Can Hear Forever

This page will work best with Netscape

Antenna Calculations

This will calculate the length of a 1 wavelength loop

Freq. in Megahertz	length of loop in feet		

This will calculate the length of a half wave dipole.

Freq. in Megahertz	length of half wave length wire in feet		

This will calculate the length of a 5/8 wave vertical with 1/4 wave radials.

Freq. in Megahertz	length of 5/8 wave vertical in feet	length of 1/4 wave radials in feet		

Freq. in Megahertz	length of L1 in feet	length of L2 and L3 in feet		

HOME

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Why I use dBi---Mostly

L. B. Cebik, W4RNL

Although current commercial practice is to specify antenna gain in dBd, I tend in my modeling work to use dBi. It is reasonable to ask why, if for no other reaosn than to understand better the other entries in this collection of notes. So I shall explain why. My object is not to get anyone to change their preferred ways, but only to make a little clearer my preference for specifying gain in dBi.

dBi

Every measure in decibels, or dB, is ultimately a relative power measurement (with some defined relative voltage and current measures having been derived from the basic power measure). Decibels are defined this way:

Relative power in dB = 10 log (base 10) P1 / P2,

where P1 and P2 are two power levels measured in the same units (e.g., watts).

Since we may pick any two powers for P1 and P2, power gain or loss is strictly relative.

However, we have found it useful to select--for specific purposes--certain baseline power levels. One such level is whatever power there might be in the radiation far field of an isotropic radiator. An isotropic radiator is a lossless dimensionless point in free space that radiates equally well in all directions. Although some say this is a absolutely theoretic concept only, Brian Beezley, K6STI, has established that a pretty good appoximation of an isotropic radiator can actually be constructed.

Using this radiator as the baseline and taking measurements at the same far field distance from the antenna, the power received from the antenna will have a certain relative level in comparison to what would have been received from the isotropic radiator. What that level is depends on the characteristics of the test antenna and the direction in which we choose to take the readings, using a full 3-dimensional sphere as the possible directions for readings. In some directions we may get more power; in others we may receive less.

One of the chief advantages of the isotropic radiator is that its field never changes, so that it functions as

an agreed upon constant against which every antenna may be measured in every direction.

Every antenna measurement referenced to a P2 that is the far field power from an isotropic radiator has a positive or negative gain in dBi.

Gain

When we casually refer to antenna gain, we do not normally mean the gain in any haphazard direction from the test antenna. Nor do we mean all the possible gain numbers we might gather from a systemmatic tour around the surface of our far field globe surrounding the antenna. Gain in every direction is important, for it defines the antenna pattern and tells us where the far field is strongest and weakest and in the middle--and by how much.

However, when we casually mention gain, we are usually interested in the direction(s) of maximum gain from a given test antenna. Then we rotate the antenna to present that antenna direction to the receiving station. For satellites, we may rotate in 3 dimensions, but for HF, 2-dimensional rotation usually suffices (unless we must somehow compensate for a bit of tricky terrain and have the wherewithal to do so).

We can specify maximum antenna gain in terms of dBi. Then we can compare maximum antenna gains from 2 or more antennas by citing their gains in dBi and merely comparing numbers.

This practice is perfectly reasonable so long as we do not make it a fetish; that is, so long as we do not use this number exclusively in our gain considerations. Horizontal beamwidth is also important and is usually defined in terms of the number of degrees between half-power or -3 dB points where the direction of maximum gain is the center. Vertical beamwidth is also important in estimating the success of a potential path. We also want to know, in conjunction with vertical beamwidth, the elevation angle of maximum radiation. Together, these numbers give a more complete picture of antenna performance in a desired direction than raw gain alone.

I have modeled an array with a 21 dBi gain figure. However, the horizntal beamwidth is only about 17 or 18 degrees wide, making it unsuited for general amatuer operation. However, we often assume that competing antennas have the same beamwidth and that all we need to know in making our selection is the maximum gain. Good practice calls for making no such assumption. Rather, before we focus on maximum gain, we should establish that all other factors are equal.

dBd: a warm but fuzzy concept

The concept of dBd was formed to capture the gain of an antenna relative to a dipole. A dipole is considered the standard basic horizontal antenna, and comparisons to it seemed to some folks to be more meaningful than comparisons to the isotropic radiator.

Unfortunately, the concept of dBd has become a cluster of concepts. Here are some of them. The notation is my own, since few folks are anxious to distinguish the individuals in the cluster.

1. dBd-I: dBd ideal compares the gain of an antenna to an ideal dipole in free space. An ideal dipole uses infinitely thin lossless wire and is resonant at the frequency of interest. In this application, the ideal dipole has a gain of about 2.15 dBi, that is, 2.15 dB over an isotropic radiator. All measurements are thus arithmetically transportable between dBi and dBd-I by adding or subtracting 2.15, as appropriate.

dBd-I is of limited utility, since my backyard dipole may have a gain of 7.15 dBi and 5 dBd. Some folks become confused by the idea that a dipole has gain over a dipole. We then have to explain that the real wire dipole has gain over the perfect dipole in free space. That rarely helps a lot. And it does not tell us until we do some arithmetic how much gain some other antenna has over a real dipole.

2. dBd-RM: dBd can be expressed as the gain over a real dipole modeled at the same height as the test antenna. For studies that are strictly modeling investigations, this measure is sometimes useful. However, it requires that we further specify the construction of the dipole in terms of element diameter and element material. Using the same material, dipole gain will vary with element diameter. It will also vary inversely with the loss of the material used for the dipole.

Both these constraints apply within the further rule of keeping the dipole resonant. If we make measurements across a ham band, we shall find that the dipole gain varies with frequency unless we reresonate it for each readout frequency. Actually, we are usually more careless than this and take one resonant reading and apply it across the band without checking. And we tend to use fairly loose standards of resonance rather than converging the results. We may more precisely say that an antenna is resonant when the reduction of feedpoint reactance results in no further changes in gain to the number of significant digits that apply to the test.

As one more qualification, we should note that dipoles and other antrennas may have different elevation angles of maximum radiation over the same type of ground. When we cite dBd-RM, we must also say whether we are giving the figure for the dipole at its angle of maximum radiation or at the angle chosen for the test antenna. To avoid confusion, it is usually better to give details about the differing elevation patterns.

3. dBd-RR: dBd can be expressed as the gain over a real dipole set in the same position as the test antenna, where both antennas are oriented for maximum gain relative to the far field receiving site of the test range. For fairness, one should specify the construction of the dipole to ensure that the materials are comparable to those of the test antenna.

However, there are a number of variables that occur within this way of handling dBd. First, range conditions vary considerably from one site to another. Second, some testers take the average of a number of readings in various directions, while others take readings along a single line defined as the best test line. Third, different ranges may use different test heights. The importance of this factor stems from the

fact that many antennas that might be tested have different elevation angles of maximum radiation than a dipole, and this variance may introduce differences in readings as we change test antennas and as we move from range to range with different test antenna heights.

Good testing and modeling protocols would specify all of the relevant factors applying to the comparisons involved. We sometimes do find these specifications. Unfortunately, we often don't. Without the specifications, comparisons in dBd-R (either M or R) are quite difficult to make. If we could only bring all antennas to a single test range with a single (large) set of dipoles and standardized conditions, we could likely establish the gain of each antenna over its standard dipole and then have truly precise comparisons among antennas. Someone has noted to me that the odds of the earth being struck by a meteor like the one which ended the reign of the dinosours are higher than the odds of the emergence of a universal test range. I really wish I could have disagreed with this individual.

For additional pitfalls of careless dBd-ing, see the next item on the Index.

Why I tend to stick to dBi

My antenna work includes the building of test models of antennas that are feasible to construct, but is devoted predominantly to modeling all sorts of antennas for all sorts of purposes. This factor alone suggests the use of a single standard for all comparisons, such as dBi. However, there are a number of reasons I tend not to use dBd except in special circumstances.

1. The relevant comparisons are not with a dipole. Very often, antenna comparisons are among antennas that do not include dipoles. In such cases, simply comparing gain figures in dBi tells us all we need to know and can know about the relative maximum gains of the antennas. For example, in comparing the gains of self-contained 1 wl wire-loop antennas, the best designs for a given purpose are the ones that are best within the group, and the group does not include a horizontal dipole.

Likewise, when contemplating whether it is worthwhile to increase the size of a Yagi for 20 meters from 3 to 4 elements, the dipole is not relevant. Rather, the relevant gain comparison is between both models and real antennas of 3 and 4 element design. The gain advantage is derived as easily from dBi as from any other system of gain numbers.

2. The relevant comparisons have only passing reference to maximum gain. Additional factors, such as elevation angle, beamwidth, etc., may be far more important than gain itself. For some applications, good antennas do not need gain relative to a dipole. But thay may require close specification of other antenna properties. Gain becomes a secondary specification for which dBi suffices nicely.

Range tests are another matter

If my work were primarily with real antennas, then dBi would become a problematical term. A test range

approximation of an isotropic radiator is an unlikely event anytime soon. Hence, antennas on test ranges must be compared to some standard, and the dipole is the most likely simple horizontally polarized candidate. If we accept this premise, then it is unreasonable to expect range testers to then correlate their results to a scheme of modeling in which the gain is converted to a value in dBi.

However, this situation makes it imperative that range testers specify a test protocol and comparison antenna for the evaluation of the procedure against good engineering practice. In many instances, the comparison antenna will not be a dipole (although amateurs often think only in terms of Yagi tests). Verticals require specification of a vertical standard of the appropriate class. Nothing substitutes for the revelation of detailed test protocols where range testing is at stake.

Conversion of the test situation to models--like any other case of modeling--is at best the most reasonable approximation we can develop. Absolute precision is unlikely in most instances. Hence, the conversion of range test results to modeling results--with the accompanying conversion of dBd-RR (with a clear specification of the standard dipole used) to dBi--is not an automatic process and is subject to varying degrees of adequacy.

Modeling of detailed test protocols, on the other hand, can yield some correlation factors among different test protocols. A model may substitute one standard dipole for another, may insert or remove test range objects affecting results, may change antenna heights with ease, etc. As a simple example, if Smith tests his antenna at 60' up and Jones tests his at 85' up, the differences in test results (all other facors being equal or equalized) can be at least tentatively resolved with effective modeling.

Whatever the prospects for such work in the future (for we have just begun to scratch the surface of effective modeling in antenna work), those who predominantly model will likely stick to dBi as the basic measure of maximum gain. Range testers will likely stick to dBd-RR, where the range test is a comparison to a dipole or other relevant standard.

Since I am mostly a modeler, I use dBi---mostly.

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How High should my Dipole be?

Dipole Antennas - the Effect of Height Above Ground

I frequently hear the question: how high should my dipole be? Or alternatively, will my dipole work well at this or that height? Unfortunately, these questions can not be answered without first stating what you want the dipole to actually DO, i.e. how you plan to operate with it. Some possible goals for a dipole might be:

1. DX work.
2. Local work: nets and rag chewing.
3. Directionality: gain in one direction, or nulls in some other direction
4. Omni-directionality.
5. Feed point impedance of 50 ohms.

As you may surmise, many of these potential goals are mutually exclusive, or at least tradeoffs.
However, once you define what you want to do with your dipole, then you can look at the radiation patterns to see if it will accomplish those goals.

I make the assumption that anyone reading this understands that DX work requires a low angle of radiation, with gain in the favored direction being desirable. Nets and rag chewing require a much higher angle of radiation and an omni-directional pattern. A null aimed in some direction may be desirable in various situations. Something close to 50 ohms impedance will aide matching and power transfer to/from coax cable.

There are many other potential goals for a dipole, but the ones I have listed are those that are most dependent on it's height above ground. Thus this discussion will not touch upon the issues of multi - banded operation, tuned open wire feeders, and the like. So, lets limit the issue at hand to: how the character of a dipole varies with its height above ground.

To investigate this problem, I have modeled a hypothetical wire dipole using the EZNEC program (from W7EL). This model is well within the verified capability of EZNEC.

For those interested in the modeling details, this dipole, named D40M, has the following specifications:
Material: #12 copper wire.
Length: 69.057 feet.
Ground Type: good (.005,13) NEC Sommerfield.
Frequency: 7.00 MHz nominal, but the comparative patterns were computed by adjusting the frequency slightly for resonance at each height.

The dipole was modeled at various heights from .05 wavelengths (7 feet) to 4 wavelengths (560 feet) above

good ground. One may argue that 560 feet is ridiculous for a 40 meter dipole, but keep in mind that the data can be scaled down to a 10 meter dipole with similar results.

The table below tabulates the results. In the first two columns, the antenna's height above ground is given in wavelengths and in feet. The next two columns show the maximum gain in the favored direction (i.e. broadside to the wire), followed by the launch angle and the -3 dB vertical beam width. The next two columns once again present the gain and launch angle / beam width, but for the axial direction (off the ends of the wire). Finally, the last 2 columns list the complex impedance at the feed point, and the actual resonance frequency at that specific height.

Height	**Height**	**Fav Dir**	**Fav Dir**	**End Dir**	**End Dir**		
			Launch		**Launch**		
Wave		**Gain**	**Angle/**	**Gain**	**Angle/**	**Feedpt**	**Res.**
Lengths	**Feet**	**(dbi)**	**Bmwidth**	**(dbi)**	**Bmwidth**	**Z**	**Freq**
4.0*	560	7.75	**4 / 4**	5.57	72 / 13		6.93
3.0	420	7.83	**5 / 5**	5.25	68 / 14	77+ j11	6.94
2.0*	280	7.80	**7 / 7**	0	39 /	75 + j12	6.95
1.5	210	7.72	**9 / 10**	-2.50	33 /	75 + j11	6.96
1.0*	140	7.64	**14 / 15**	-11.00	20 /	74 + j08	6.96
.9	126	7.03	**16 / 17**	-8.30	22 /	85 + j13	6.94
.8	112	7.16	**18 / 19**	-6.40	25 /	84 + j26	6.88
.7*	98	7.95	**20 / 22**	-4.50	30 /	70 + j30	6.88
.6	84	8.35	**23 / 26**	-1.95	40 /	60 + j16	6.94
.5*	70	7.45	**28 / 33**	-0.51	43 / 33	71 - j00	7.00
.4	56	6.06	**35 / 47**	1.30	59 / 102	93 + j04	6.98
.3*	42	5.59	**50 / 137**	4.71	90 / 80	100 + j32	6.86
.2	28	6.70	**90 / 118**	6.70	90 / 67	71 + j56	6.77
.1*	14	8.21	**90 / 103**	8.21	90 / 66	23 + j39	6.84
.05	7	9.61	**90 / 99**	9.60	90 / 72	7 + j12	6.95

***** Elevation plots shown below

Analysis, Favored Direction:

The first thing to notice is that the gain in the favored (broadside) direction varies very little with height. The important change in the broadside pattern occurs in the launch angle of the primary lobe. As the antenna

moves closer to the ground, the launch angle of radiation gets higher and the -3 dB vertical beam width becomes broader. Note that below the benchmark height of ½ wavelength, the launch angle increases rapidly. Once the dipole is lowered to 0.3 wavelengths, most of the radiation goes in a vertical direction. This explains the frequently heard "rule" that a dipole must be at least ½ wavelength high to work. The seeming anomaly with the beam width below 0.4 wavelengths is easier to understand by viewing the plots shown below.

Analysis, End-Fire Direction:

One frequently sees a dipole azimuth pattern depicting a very sharp null off of the ends of a dipole. While technically accurate, this can be very misleading as the table above shows, and is a result of trying to depict a 3 dimensional pattern in 2 dimensions. This often seen null is only evident at the same launch angle as the maximum broadside gain. Of major significance is the large amount of gain off the ends at higher launch angles. Due to multiple lobes forming above ½ wavelength, this is not easily shown in tabular form. I have arbitrarily chosen to list gain and launch angle for the secondary lobe with the lowest launch angle, but recognize that there is frequently a stronger primary lobe at higher angles. Consult the plots below for a better visualization.

Analysis, Feed Point Impedance:

The reference antenna length was chosen to resonate at the ½ wavelength height. As expected, the feed point impedance oscillates significantly as the height changes from our reference point. Thus we verify the old adage that you must trim the dipole to fit your particular QTH (height being very important). The corresponding resonant frequency for each height is shown in the last column for reference, since complex impedance's may be of less practical importance to some.

So, How High should the dipole be to work well?

Now we are back to looking at what we want the dipole to achieve.

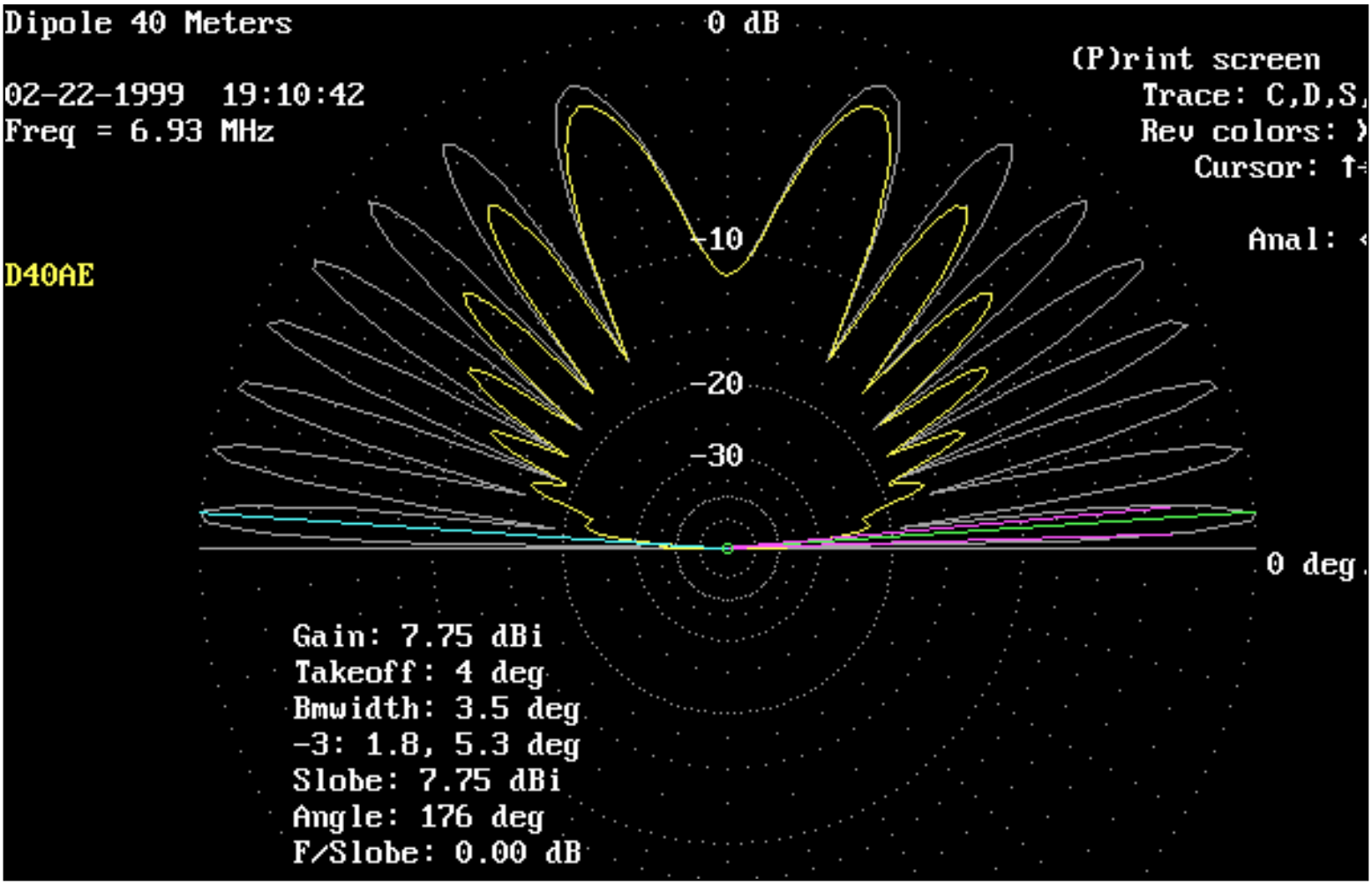
For DX work, higher placement is warranted, since more power concentrated between 5 and 15 degrees is reported to be of major benefit. Heights around one wavelength are necessary to get the broadside lobe to launch in this range. However, higher may not always be better. Pay careful attention to the magnitude of secondary lobes in the broadside direction, as well as high angle radiation off the ends. Some heights would appear better than others due to concerns with nulling out local QRM. A complete discussion of of this aspect is beyond the scope of this article, but may be investigated at a later date.

For local work, lower heights appear to be more beneficial. Note especially how omni-directional our dipole becomes at lower heights. Below 0.4 wavelengths, there is less than 1 dB of attenuation in the end fire direction, which suggests a height between 0.4 and 0.3 might be an ideal compromise for local nets and rag chewing.

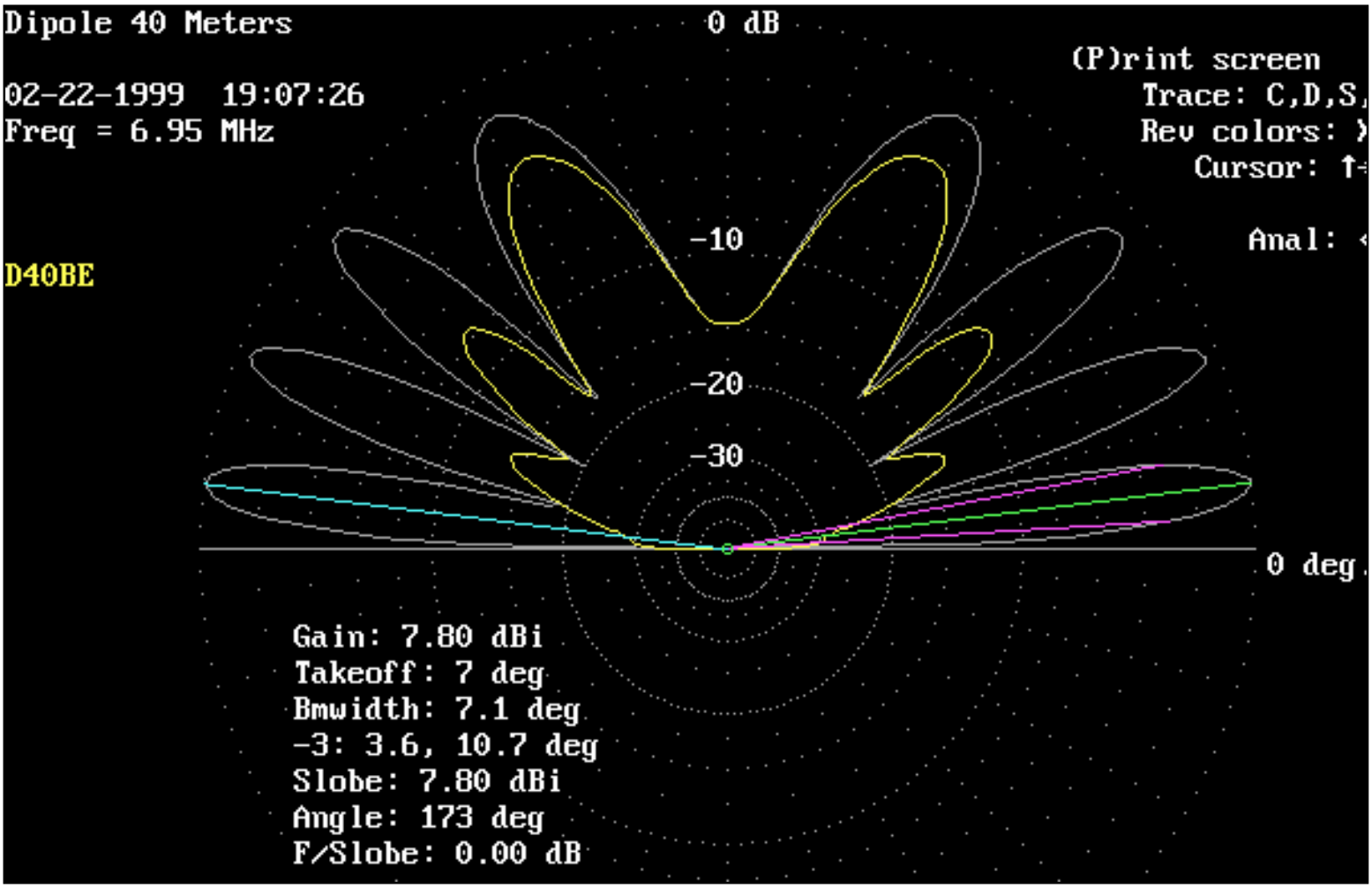
Feed point impedance and matching does not seem to be of major concern except at very low heights. The effect of height on 2:1 SWR bandwidth was not investigated.

Radiation Plots

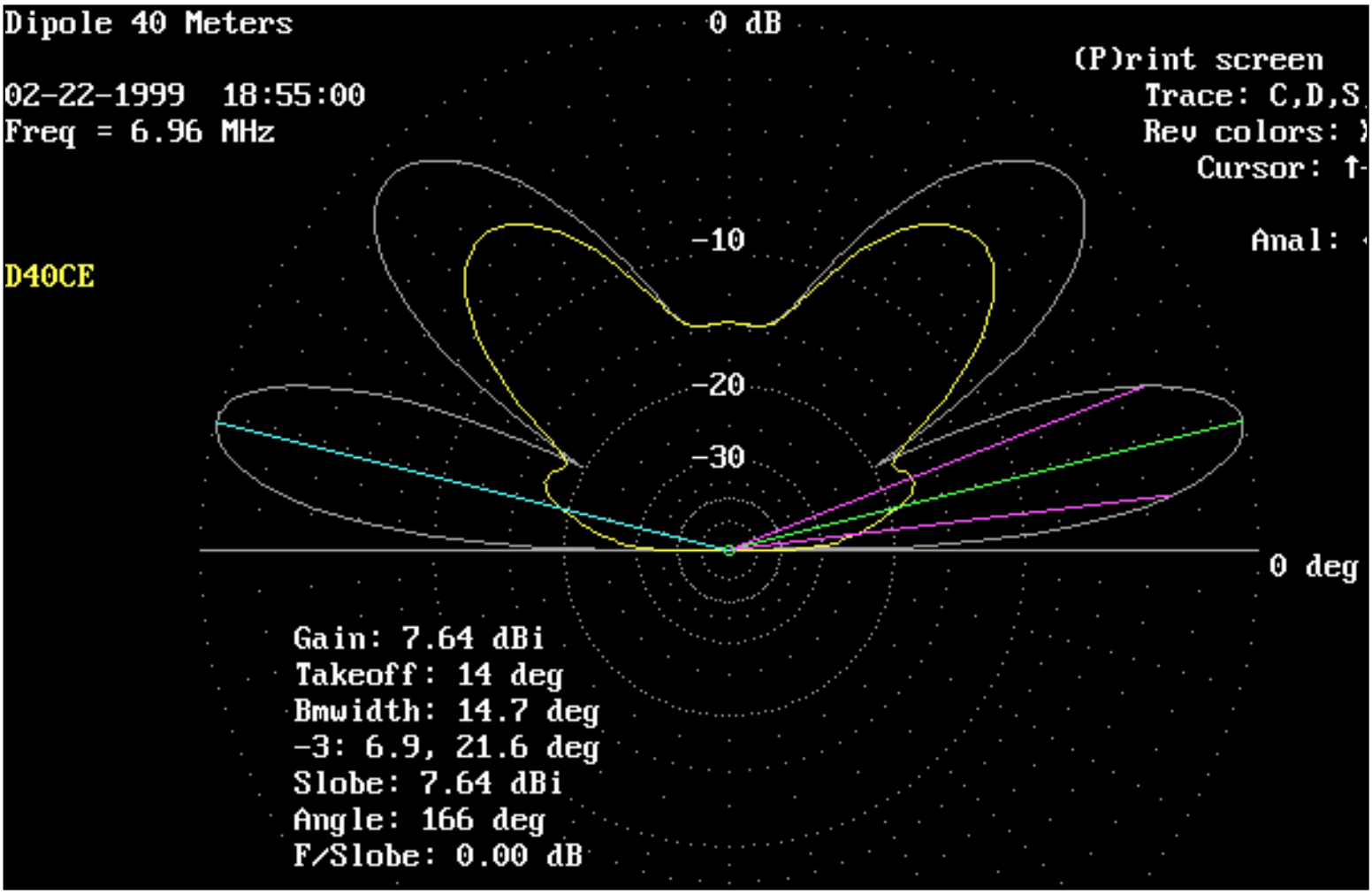
Our first plot shows the elevation pattern in both the broadside (white trace) and end fire (yellow trace) directions for a height of 4 wavelengths, or 560 feet. Note the substantial high angle radiation in all directions, in addition to the excellent low angle lobes.



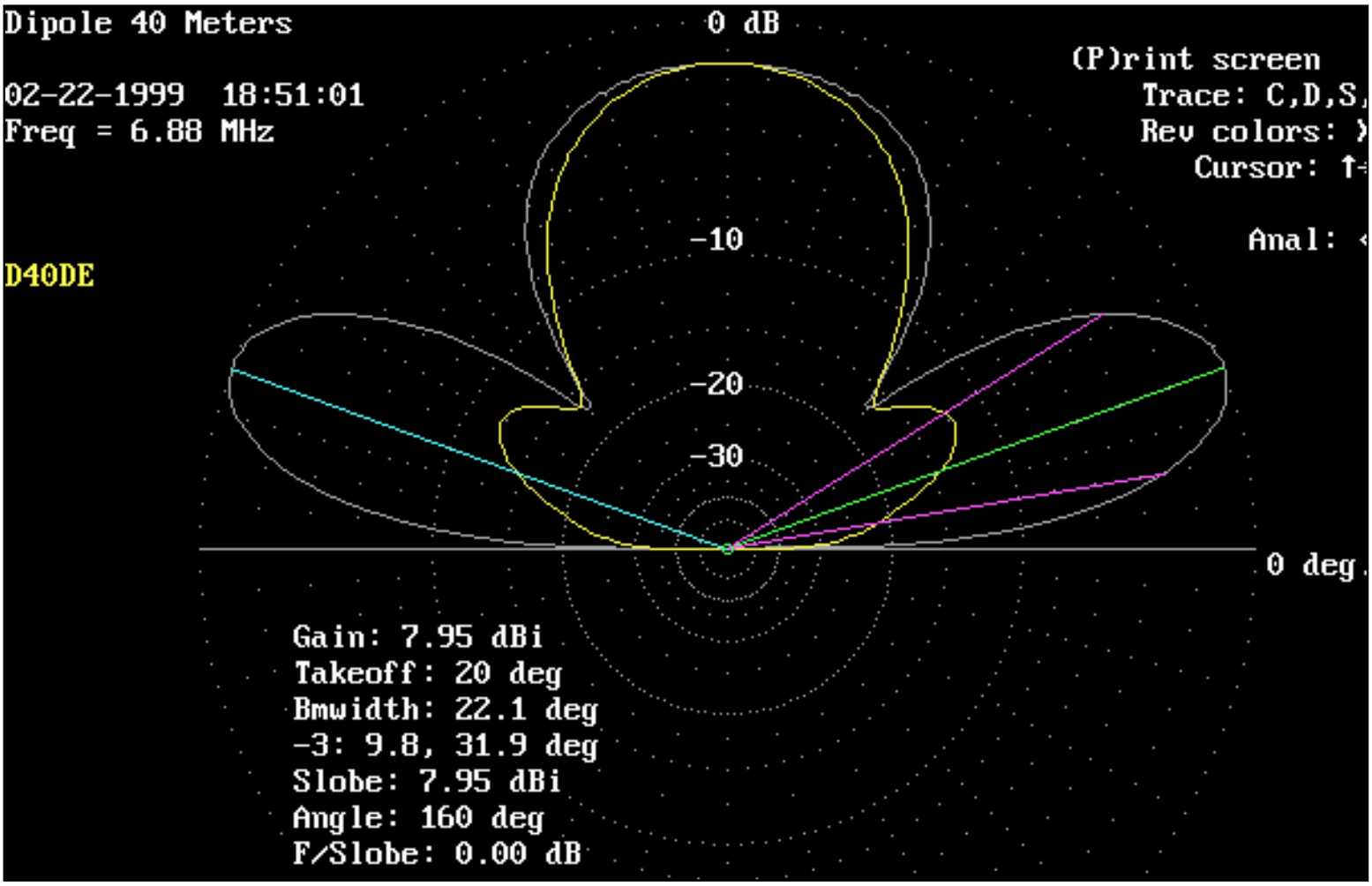
This plot shows the elevation patterns at a height of 2 wavelengths, or 280 feet.
White trace is broadside. Yellow trace is axial (off the ends).
Still Lots of high angle radiation.



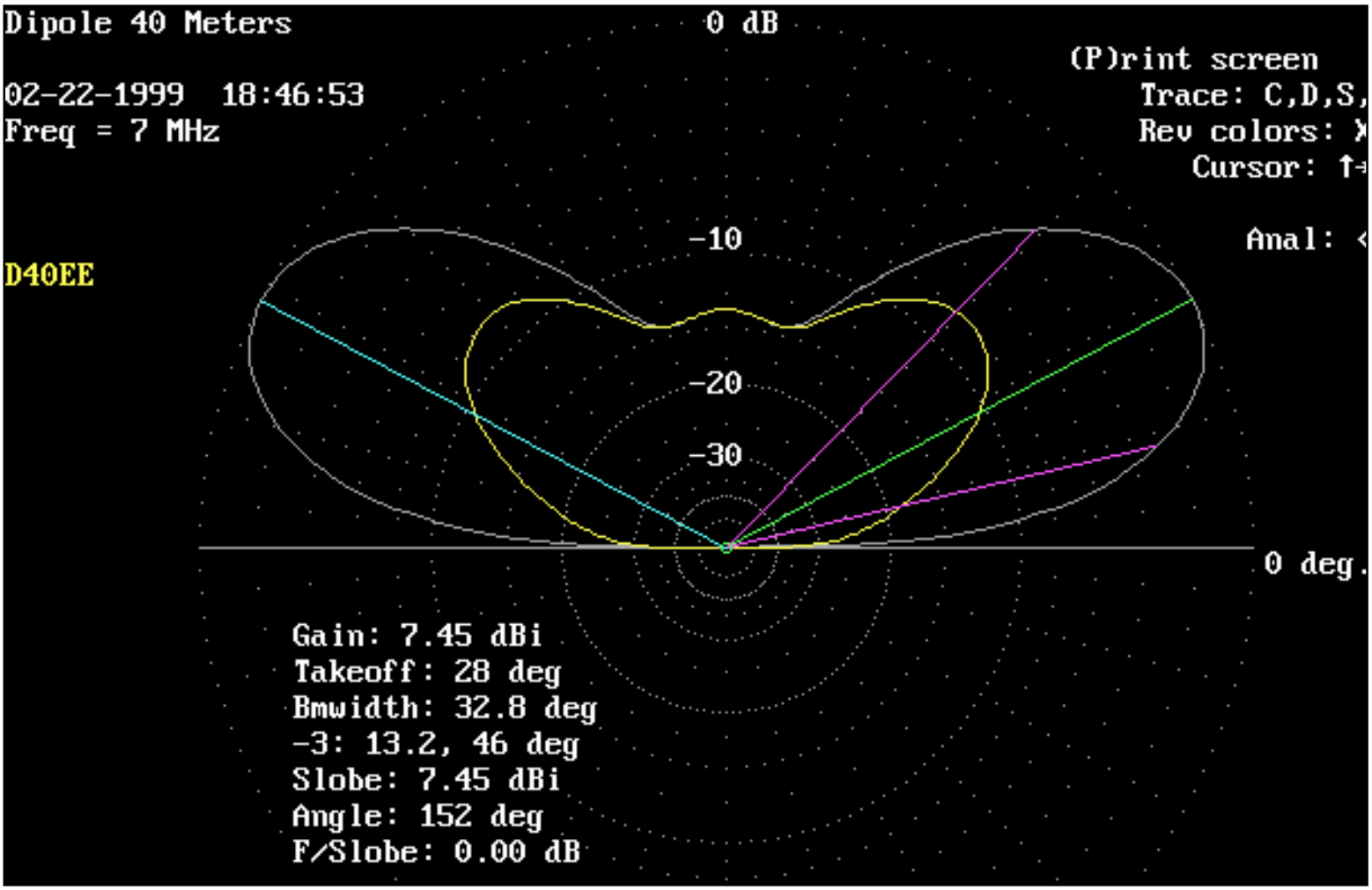
This plot shows the elevation patterns at a height of 1 wavelength, or 140 feet.
White trace is broadside. Yellow trace is axial (off the ends).
The secondary lobe is down to 47 degrees, but the primary lobe is up to 14 degrees.



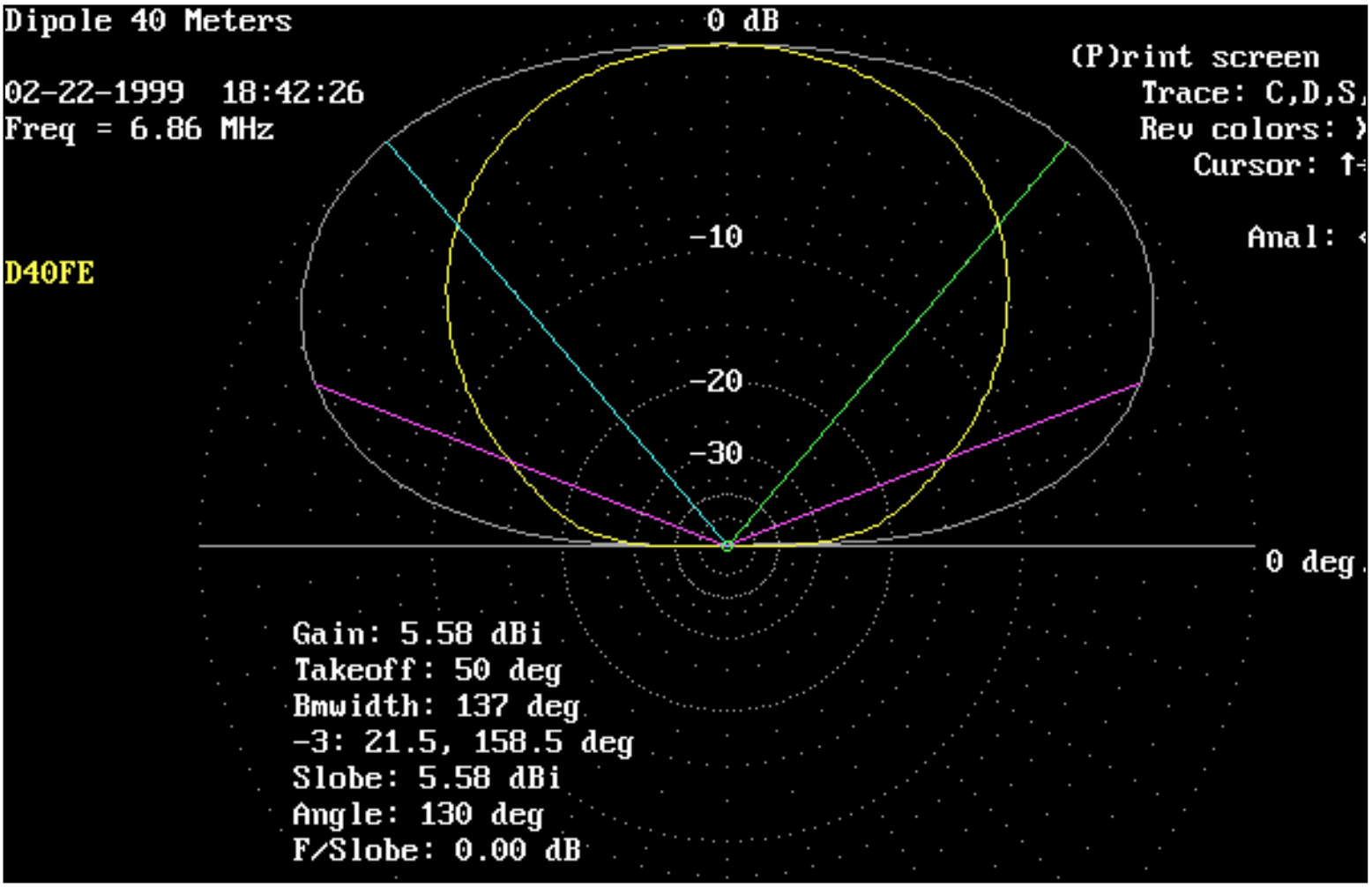
This plot shows the elevation patterns at a height of 0.7 wavelength, or 98 feet.
White trace is broadside. Yellow trace is axial (off the ends).
The primary lobe is up to 20 degrees.
Note the large vertical lobe which has appeared!



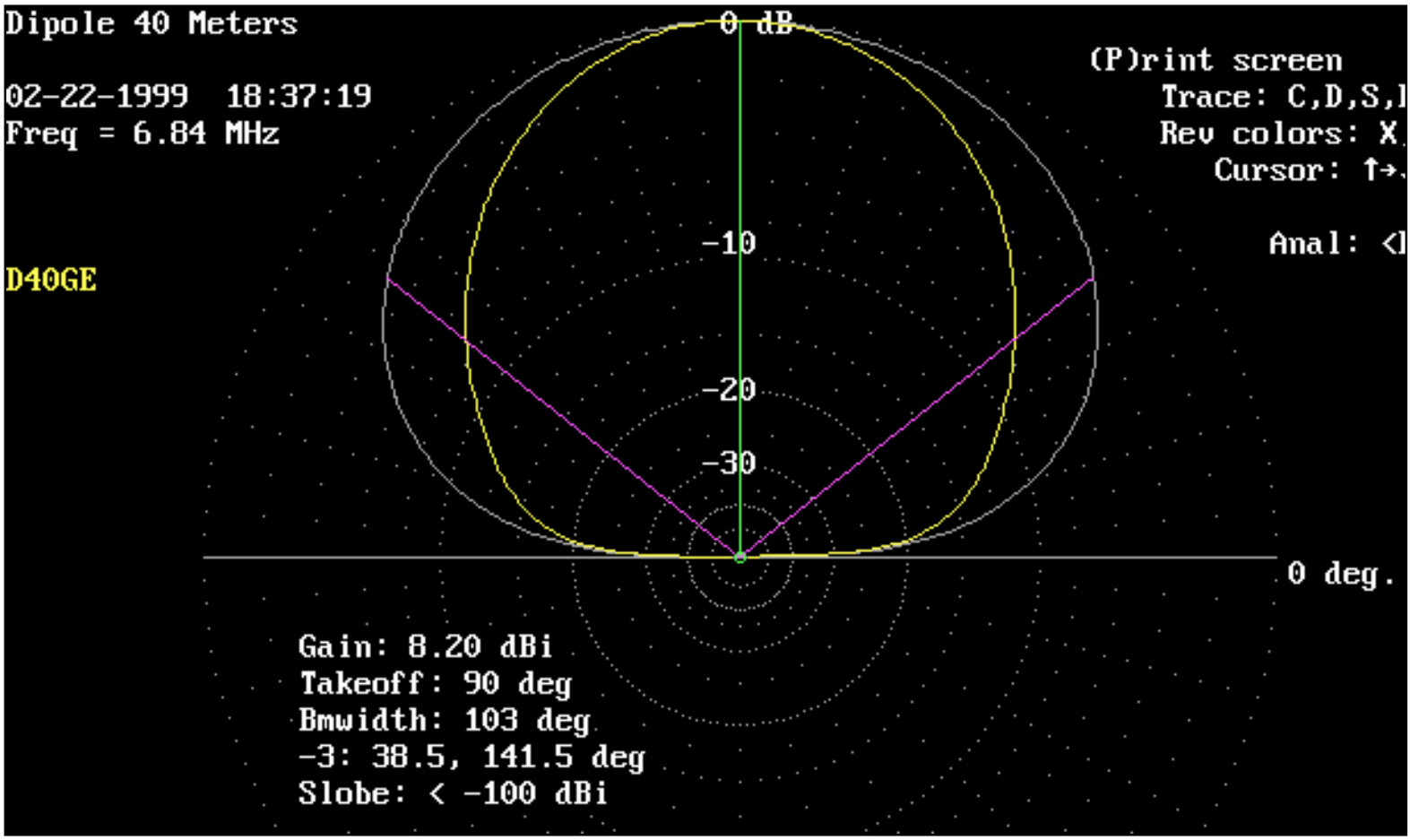
This plot shows the elevation patterns at a height of 1/2 wavelength, or 70 feet.
White trace is broadside. Yellow trace is axial (off the ends).
Now that's a classical dipole pattern!



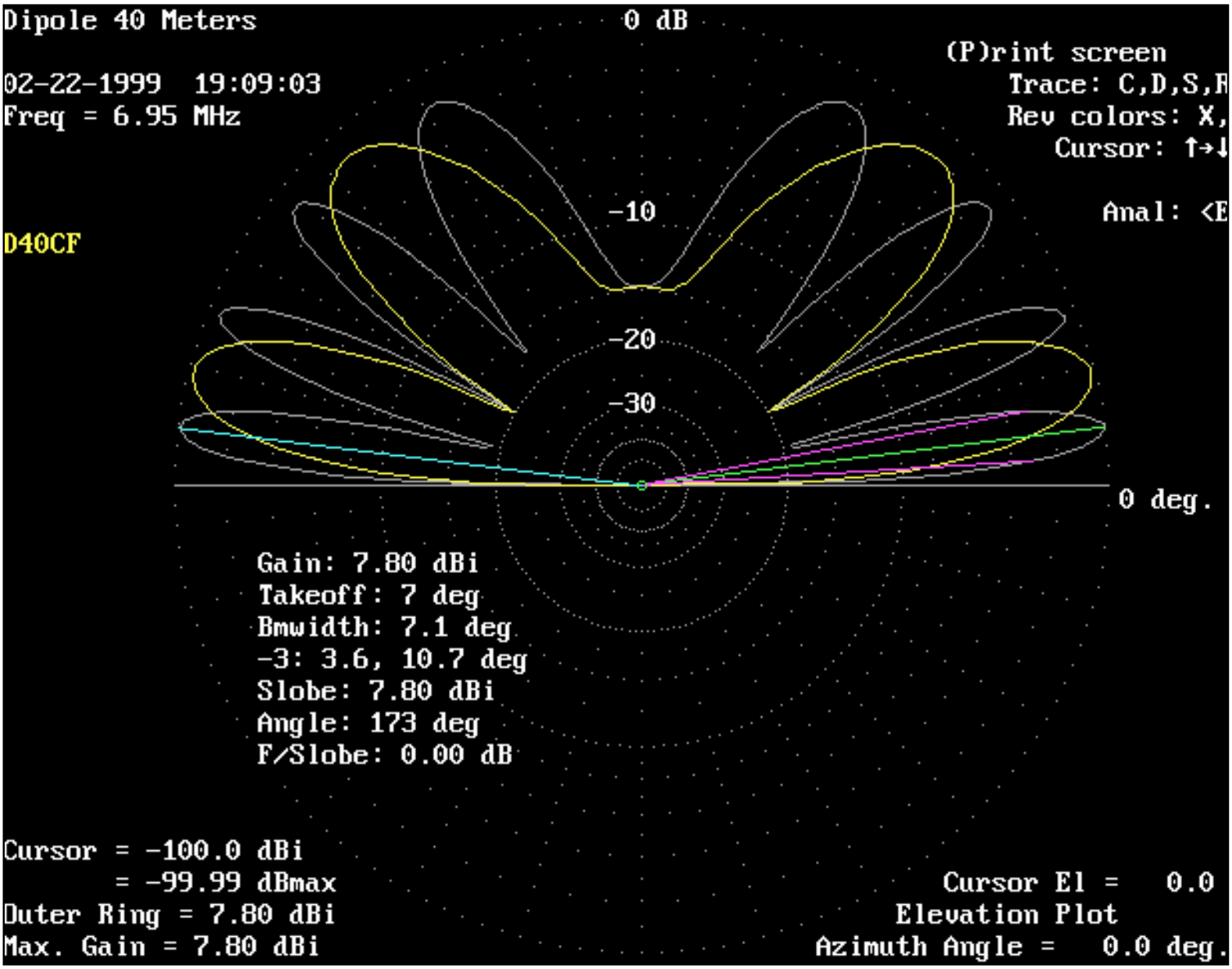
This plot shows the elevation patterns at a height of 0.3 wavelength, or 42 feet.
White trace is broadside. Yellow trace is axial (off the ends).
We are entering the "skywarmer" mode here.



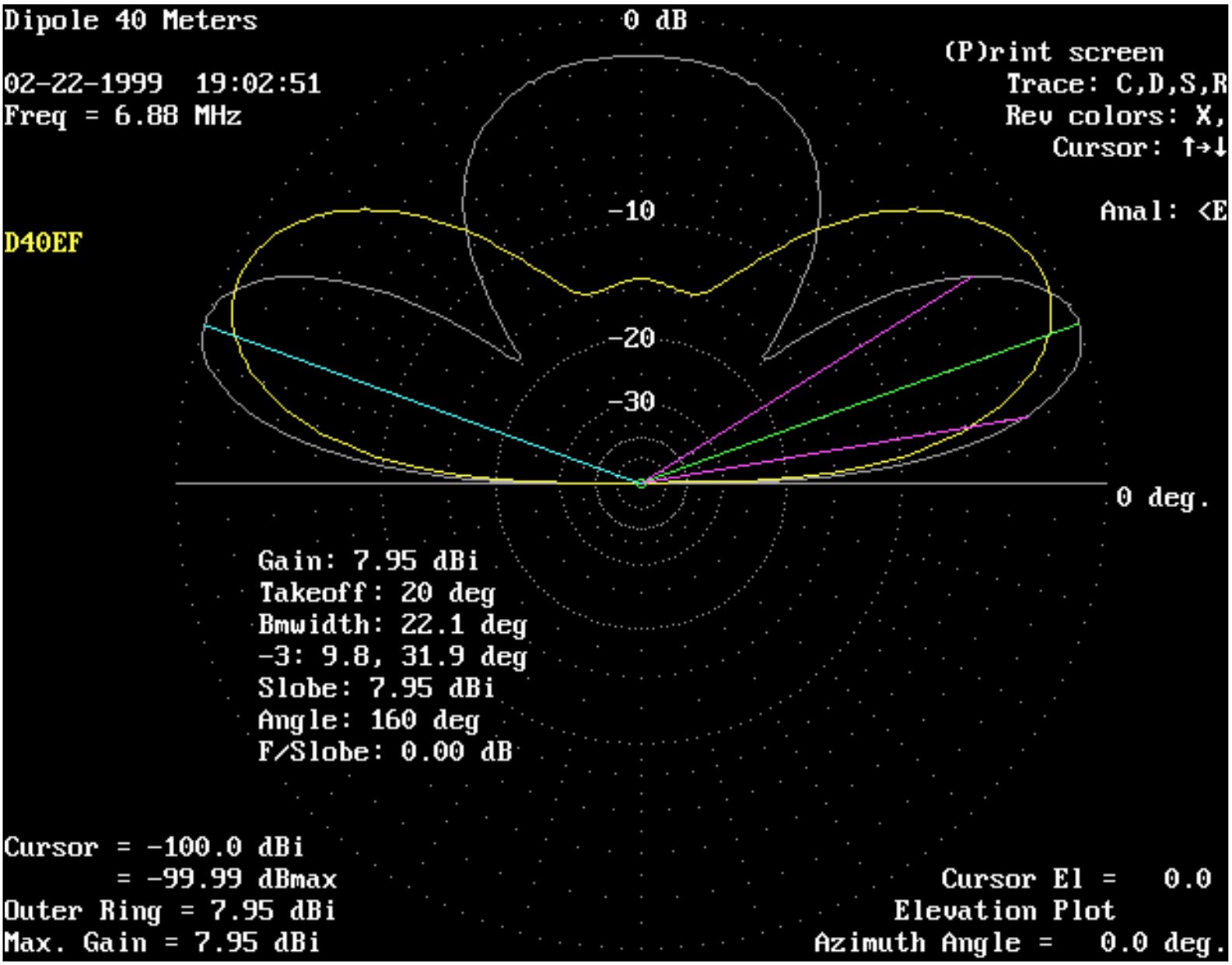
This plot shows the elevation patterns at a height of 0.1 wavelength, or 14 feet.
White trace is broadside. Yellow trace is axial (off the ends).



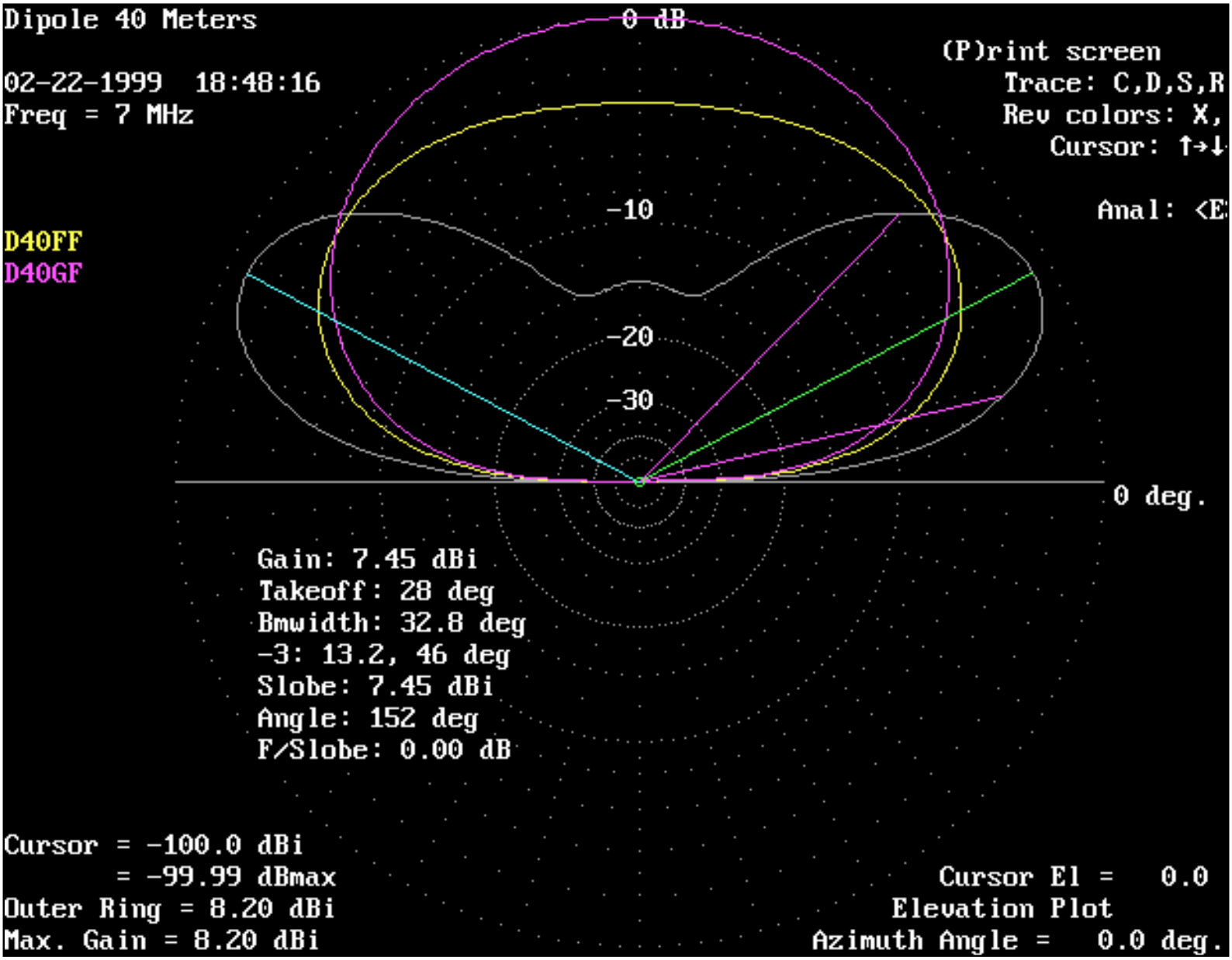
This plot shows comparative elevation patterns, in the broadside direction only,
White trace is for 2 wavelengths, 280 feet.
Yellow trace is for 1 wavelengths, 140 feet.



This plot shows comparative elevation patterns, in the broadside direction only.
White trace is for 0.7 wavelengths, 98 feet.
Yellow trace is for 0.5 wavelengths, 70 feet.



This plot shows comparative elevation patterns, in the broadside direction only.
White trace is for 0.5 wavelengths, 70 feet.
Yellow trace is for 0.3 wavelengths, 42 feet.
Pink trace is for 0.1 wavelengths, 14 feet.



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Dipole Height

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Antenna Tuner Operation

I'm a little confused about how antenna tuners function in terms of reducing antenna system SWR. Is it true that an antenna tuner does not really 'tune' the antenna? And if an antenna is cut to resonance already, what good is a tuner?

Yes, it's true-an antenna tuner doesn't really tune your antenna in the strict sense of the word. It does not, for example, adjust the lengths of your antenna elements, their heights above ground and so on. What an antenna tuner or transmatch does do, however, is transform the impedance at the feed line input to a value that your transceiver can handle (typically 50 Ohm-see Figure 1). When thinking about antenna tuners and SWR, it's important to remember that the tuner has no effect whatsoever on the SWR between itself and the antenna. It's the SWR between the tuner and the transceiver that changes.

In practical terms, all a tuner does is act as a kind of adjustable impedance transformer between the antenna system and the radio. It takes whatever impedance the antenna system presents and attempts to convert it to 50 Ohm--or something reasonably close to that value--for the transceiver. When the transceiver "sees" a 50 Ohm impedance, it is able to load its maximum RF output into the system. That power is transferred through the antenna tuner, to the feed line and, ultimately, to the antenna--minus any losses incurred along the way.

These losses are the reason that the highest efficiency feed-line for each individual case is desirable and why some amateurs use ladder line.

What is the correct way to tune an antenna tuner?

Most antenna tuners have an inductance rotary switch and two capacitors. The capacitors are often labeled ANTENNA and

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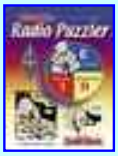

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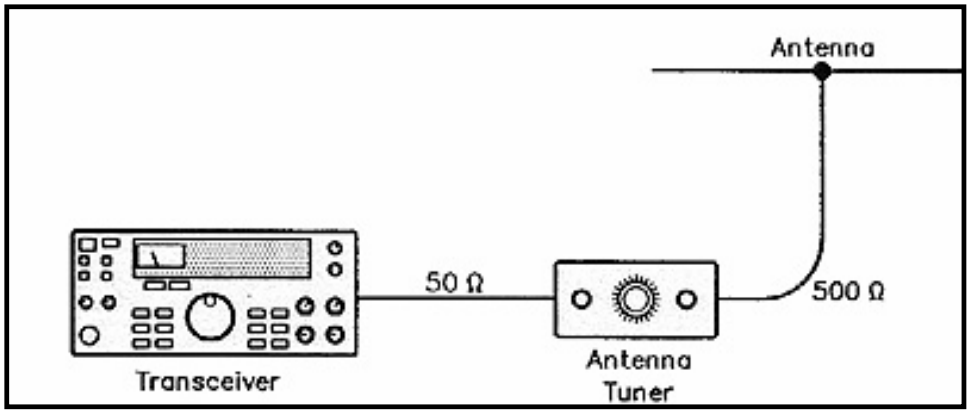

(Figure) (QST February 2000, p. 49)

TRANSMITTER. In some antenna tuners the inductance switch is replaced with a continuously variable inductance, popularly known as a roller inductor.

Let's assume you're using a tuner with an inductance switch, because they are the most common. Place both capacitor controls at their mid-range positions. Don't trust the knob markers if this is your first experience with the tuner; remove the cover and turn the knobs until the moving capacitor plates are only half meshed with the stationary plates. If the knobs are pointing to half scale, consider yourself lucky. If not, loosen their Allen nuts and rotate the knobs so that they point to mid scale. Replace the tuner cover and you're ready to go.

Turn the radio on and, with the ANTENNA and TRANSMITTER controls at mid scale, crank the inductance switch until you hear the loudest noise or signals coming into your radio. Then, rotate the ANTENNA and TRANSMITTER controls until you get to the absolutely loudest noise or signal level on the radio. This should be close to your best tuning spot.

With your rig set to low power, send an ID then transmit a continuous carrier while you tweak the ANTENNA and TRANSMITTER controls for the lowest reflected power reading with the highest output power as read on the SWR meter. You may find that you have to vary the position of the inductance switch a position or two to get your best match. Be gentle to your radio; keep the key-down periods as short as possible.

Depending on the impedance at the antenna input (and the overall design of the tuner) you may not be able to obtain a flat 1:1 SWR on all frequencies and bands.

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Do You Need an Antenna Tuner?

Maybe yes, maybe no. It all depends on the type of antenna and feed line you're using.

By Steve Ford, WB8IMY
Assistant Technical Editor

There is a great cloud of mythology surrounding antenna tuners, particularly when the conversation turns to what they can and cannot do. Make no mistake, they are useful devices in the right applications. The trick is deciding whether you need one!

When Rigs and Antenna Systems Disagree

Every antenna has an *impedance* expressed in *ohms*. The same is true of the feed line you use to connect your transceiver to the antenna. Impedance sounds like a complicated beast and, to a certain extent, it is. In simplest terms, it is a combination of *inductive reactance*, *capacitive reactance* and garden-variety *resistance*.

It's probably best to avoid a long discussion about the meaning of reactance. This is "New Ham Companion," not the *Proceedings of the IEEE*. If I had several more pages to devote to this article, I'd be more than happy to bore you to tears with reactance theory. For our purposes, think of reactance as opposition to the flow of an ac signal in a circuit. In this case, the ac is the RF generated by your transceiver and the circuit is your antenna system. File this idea away for the moment. We'll come back to it later.

Meanwhile, back at the radio ranch....

The impedance of the antenna depends on a number of factors, including the length, operating frequency, height above ground, proximity of metal objects and even weather conditions (such as ice on the antenna). The impedance of the feed line depends on how the cable is constructed.

Your feed line does more than simply connect your radio to your antenna. It acts as an impedance *transformer*. That is, the impedance of your antenna is transformed by the feed line into the value your radio "sees" when you connect it to the cable. This *system impedance* acts as a *load* for the energy created by your radio—just like a light bulb is a load for the energy supplied by a battery.

Most amateur transceivers are designed to work with a load impedance of 50 ohms. When your radio sees an impedance of 50 ohms, or something close to it, you're on easy street. You press the mike switch, close the CW key or type on your keyboard and all is right with the world.

But what happens when the impedance *isn't* 50 ohms? Now you have a situation known as a *mismatch*.

When a mismatch exists, a certain portion of the power generated by your radio is *reflected*—like light is reflected by a mirror. This reflected power comes shooting back down the cable to your radio. When it reaches the radio, it is reflected back toward the antenna. The reflected power combines with the *forward* power being generated at the radio to create *standing waves* in the feed line.

By using a *standing-wave-ratio* (SWR) meter, you can measure both the forward and reflected power. A 1:1 SWR reading indicates that no power is being reflected back to your radio. This is good. On the other hand, an SWR of 3:1 or more means that a substantial amount of power is being reflected. This is usually bad. (Don't you love these simple concepts?)

A high SWR can cause considerable RF voltages to develop in the feed line and in the output circuits of your radio. This is a dangerous condition for your rig—especially if it is a modern solid-state transceiver. To prevent this, many radios manufactured within the past 10 years include SWR protection circuits. When the SWR gets too high, these circuits automatically reduce the output power or, in some cases, shut down the transceiver altogether (see Fig 1). Older tube radios are much more forgiving, but even they can be damaged when operated under high SWR conditions.

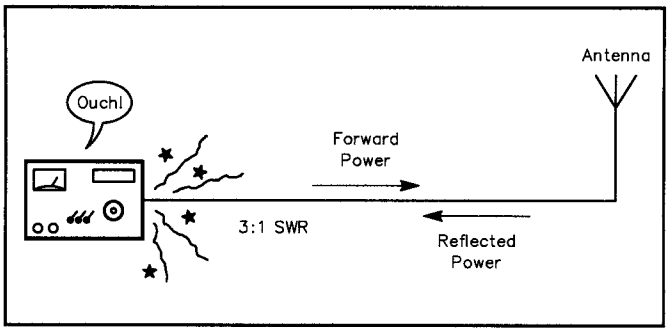


Fig 1—Most transceivers are designed to expect an antenna system impedance of 50 ohms. When the antenna impedance is something other than 50 ohms, a transmission line mismatch occurs and a portion of the RF power is reflected back to the radio. *Standing waves* are created in the feed line and high RF voltages can develop. When the *standing-wave ratio* becomes higher than 3:1, your transceiver may be damaged.

If your antenna system presents a serious mismatch to your radio, what can you do? If you connect your transceiver directly, the protection circuitry will drop your output like a rock. Worse yet, you may find yourself on the receiving end of an expensive repair bill. You need to provide a 50-ohm load for your transceiver—regardless of what is really present. One way to accomplish this is by using an antenna tuner.

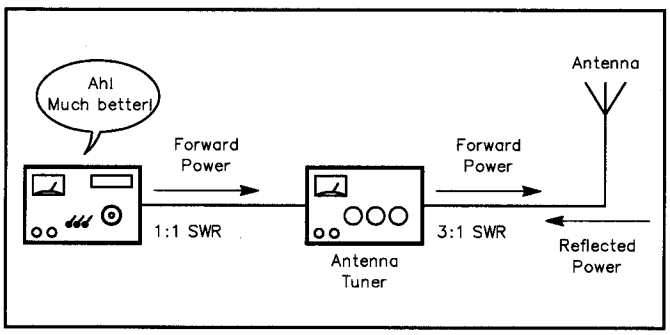


Fig 2—By using an antenna tuner, you can adjust the impedance your transceiver "sees" to a hospitable 50 ohms. The antenna mismatch to the line still exists, but the tuner protects your radio from the RF voltages while allowing it to develop its maximum output.

How Does an Antenna Tuner Work?

In its most basic form, an antenna tuner is simply a network of variable inductors (coils) and capacitors. By adjusting the coils and capacitors, you counterbalance and cancel the effects of the inductive and capacitive reactance at the *transceiver* end of the feed line. (*Now* you know why I bothered bringing up the subject of reactance in the first place!) As the reactances are canceled, the impedance at the transceiver is transformed to 50 ohms (see Fig 2).

As far as your transceiver is concerned, the load impedance is matched and it's free to dump all of its power into the antenna system.

I bet a number of you are saying to yourselves, "*Wait a minute!* The impedance at the transceiver side of the tuner is 50 ohms, but it's still some other value at the antenna side. All you've done is shift the mismatch problem from the transceiver to the tuner!"

You're right. The mismatch still exists, but now it's at the output of the antenna tuner instead of the transceiver. By using the tuner, we're protecting the radio while still allowing it to develop maximum output. If the tuner is well designed, it should be able to handle the RF voltages and currents caused by the high SWR.

Of course, the reflected power is still bouncing back and forth between the antenna tuner and the antenna. Some of this power is lost in the feed line. If you're using low-loss feed line, however, most of it is radiated at your antenna. In the meantime, your transceiver is happy and you're happy. Who could ask for more?

Use an Antenna Tuner if...

...you want to feed your antenna with open-wire line.

Open-wire line (or *ladder line*) offers extremely low loss at HF frequencies (much better than coaxial cable). One problem is that open-wire line is *balanced* while your transceiver output is *unbalanced*. You need to use an antenna tuner with a built-in *balun* to form a bridge between the balanced line and the unbalanced output of your radio. A balun is a type of transformer that converts balanced feed lines to unbalanced, or vice versa. (***BAL***anced to ***UN***balanced. Get it?) Most antenna tuners use 4:1 baluns that also convert the impedance of open-wire feed lines to a value that the tuner can handle.

...you want to operate your antenna on bands other than those it was designed for.

When you attempt to use, say, a 40-meter dipole on 10 meters, a big mismatch will develop, along with a high SWR. By using an antenna tuner, you may be able to create a 1:1 SWR at your transceiver. (I say "may" because the mismatch can sometimes be so great that it is beyond the capability of your tuner to handle.) The high SWR may cause substantial loss in a coaxial feed line, but at least you'll radiate *some* power at the antenna.

...your antenna has a narrow *SWR bandwidth* on some bands.

Some types of multiband antennas do not offer low SWRs from one end of each band to the other. There is usually a range—expressed in kilohertz—where an SWR below 2:1 can be achieved. For example, a multiband trap dipole may offer an SWR of 2:1 or less from 3600 to 3800 kHz. That's an SWR bandwidth of 200 kHz. If you try to operate above 3800 kHz or below 3600 kHz, you'll encounter an SWR higher than 2:1 and your radio may become displeased. With an antenna tuner, you can operate outside the SWR bandwidth and still load the full output of your radio into the antenna system.

Don't Bother with an Antenna Tuner if...

...your SWR is 1.5:1 or less at the frequencies you operate most often.

An SWR of 1.5:1 or less is not serious and does not require the assistance of an antenna tuner. Most modern rigs will tolerate a 1.5:1 SWR just fine. In fact, many will be happy at an SWR of 2:1. If you are using a good-quality feed line, the loss caused by an SWR of 1.5:1 or even 2:1 isn't enough to worry about at HF frequencies. Many hams are obsessed with providing an absolute 1:1 SWR for their radios at all times. Apparently they also have money to burn!

...you have a high SWR at VHF or UHF frequencies.

VHF/UHF antenna tuners are available, but my advice is to save your money. Remember that an antenna tuner massages the antenna system impedance *at the transceiver*. The mismatch still exists and the SWR is still high at the antenna side of the tuner. Even the best coaxial cables have significant losses at VHF and UHF when the SWR is high. A VHF/UHF antenna tuner will make your radio happy, but most of its power will never make it to the antenna. The best approach is to correct the mismatch at the antenna by adjusting whatever tuning mechanism it provides. If the antenna cannot be tuned, check the cable for defects and make sure you've installed the antenna properly.

...you're interfering with TVs, telephones and other appliances in your neighborhood.

Despite what you may have heard, an antenna tuner will not necessarily cure your interference problems. It's true that most antenna tuners will reduce the level of *harmonic radiation* (signals your radio generates in addition to the ones you want), and if the interference is being caused by harmonics, a tuner may help. Most interference, however, is caused by RF energy that's picked up indirectly by cables or wires, or directly by the device itself. By using an antenna tuner, you'll probably radiate more energy at the antenna than you did before. That may make your interference problem worse!

Looking for Mr Goodtuner

So, you've decided that you need an antenna tuner after all. Antenna tuners come in all shapes and sizes. What features should you consider?

❑ **A built-in SWR meter**—An SWR meter of some type is a must if you want to use an antenna tuner. When adjusting your tuner, you need to keep your eye on the *reflected power* indicator. Your goal is to reduce the reflected power to zero—or at least as close as you can get. When the reflected power is zero, the SWR is 1:1 at your transceiver.

Many tuners feature built-in meters. If not, you can purchase one separately. Your radio may even have its own SWR meter.

❑ **A roller or tapped inductor**—More expensive tuners feature a variable coil called a *roller inductor*. As you turn the front-panel inductor knob, the coil inside the tuner rotates. A metal wheel rolls along the coil like a train on a railroad track. As the wheel moves along the coil, the inductance increases or decreases.

Less expensive tuners do not use roller inductors. Instead, there is a coil with wires attached at various points. On the front panel, a rotary switch selects the wires. According to how the inductor is wired in the circuit, selecting one *tap* or another

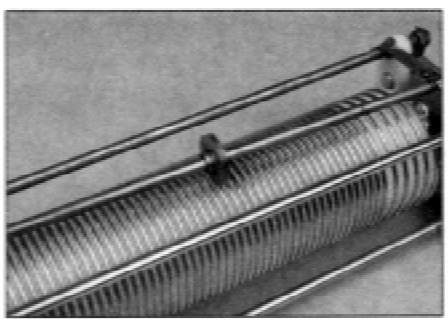


Fig 3—This is a typical roller inductor. Notice the wheel that rolls along the coil windings. As the wheel moves, the inductance changes.

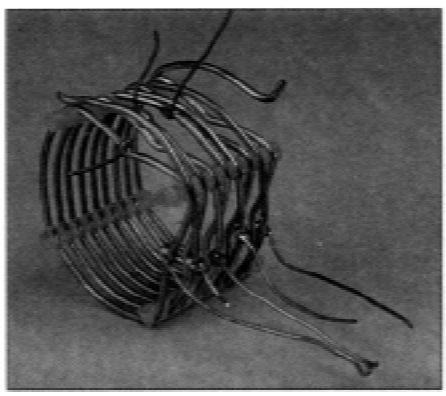


Fig 4—Tapped inductors have wires attached at various points. By selecting a particular wire, you get a fixed amount of inductance.

varies the inductance. This is known as a *tapped inductor.*

There are advantages and disadvantages to both approaches. Roller inductors offer the best tuning performance, but they are subject to the woes of mechanical wear and tear. For example, if corrosion builds up on the wheel or the coil windings, the electrical quality of the connection will deteriorate. Roller inductors are also cumbersome to use. You may have to twist the control many times when moving from one band to another.

Tapped inductors are easy to use and free of mechanical problems (unless the switches get dirty). However, you may find that they restrict the operating range of your tuner. When you turn the switch, you select a *fixed* amount of inductance. You can't easily change it to tune a particularly difficult mismatch situation.

❑ **A built-in balun**—If you intend to use an open-wire feed line, buy a tuner with a built-in 4:1 balun. These baluns often dissipate quite a bit of heat, so always choose a large balun over a small one.

❑ **Multiple antenna capability and dummy loads**—Some tuners offer the ability to connect more than one antenna. This is handy in all sorts of applications. Let's say you have a vertical antenna for 40-10 meters and a wire dipole for 80 meters. You can connect both feed lines to your tuner and easily switch between them.

Built-in dummy loads are convenient, but not necessary. A dummy load is a resistor (or group of resistors) that absorbs the output of your transceiver while allowing very little energy to radiate. It's used for making transmitter adjustments and other tests. If your tuner lacks a dummy load, you can purchase one separately.

❑ **Automatic operation**—Most transceiver manufacturers offer *automatic* antenna tuners. These tuners are usually built in the radio itself, or they're offered separately. Automatic tuners are convenient when you need to change bands or frequencies quickly. You simply push a button and your tuner adjusts its coils and capacitors to achieve the lowest SWR. Some automatic tuners sense when you've changed frequency and will readjust immediately! (You don't have to lift a finger.)

Automatic antenna tuners are expensive and their tuning range is limited. If your operating style requires you to jump from band to band rapidly (contesting is one scenario), consider an automatic tuner. Otherwise, conserve your cash and invest in a manual tuner.

A Word About Power Ratings

If your transceiver produces only 50 or 100 watts of power, a 200- or 300-watt tuner should do the trick, right? Well...yes and no. Remember what we said about mismatches causing high RF voltages in the tuner? If you're trying to use your tuner in a high-SWR situation, the RF voltages at the tuner may cause an unpleasant phenomenon known as *arcing*. That's when the RF energy literally jumps the gaps between the capacitor plates or coil windings. When your tuner arcs, you'll usually hear a snapping or buzzing noise. The reflected power meter will fluctuate wildly. Interference to your TV and other devices will increase dramatically. You may even see brilliant flashes of light inside your tuner!

Arcing is obviously bad news for your tuner. It's your tuner's way of saying, "Stop! I can't handle this mismatch!" There are two cures for arcing: reduce your output until it stops, or get a tuner with a higher power rating.

High-power tuners use large capacitors and coils. The gaps between the plates and windings are greater, making it more difficult for an arc to occur. If you can afford it, you're always better off buying a tuner with a 1.5 kW rating or better. (The exception is QRP operating where you're running low power levels.) A hefty tuner costs more, but it will serve you well in the long run.

Buy or Build?

As you comb through the advertising pages of *QST*, you'll see many new antenna tuners for sale. The prices are often reasonable and the quality is usually good. Keep your eyes open for used tuners, too. If an old tuner is in decent condition, it's every bit as usable as a new one.

If you like to build things, however, consider an antenna tuner as your next project. Antenna tuners are relatively easy to construct. You can find capacitors and coils at hamfest flea markets at low prices. Even roller inductors—the most expensive part of a roller-inductor tuner—can be found for less than $40 if you look carefully.

Your chances of success with an antenna tuner project are excellent. You have to try pretty hard to build one poorly! Best of all, you'll have the satisfaction of using a piece of equipment that you've put together yourself. The *ARRL Handbook* offers several tuner designs you can try. Heat up your soldering iron and go to it! QST

Radio Tips:

Log it or Lose it

Back in the old days (not all *that* long ago!), the FCC required every amateur to keep a detailed station log. (In addition to regular QSOs, hams even had to log unfruitful CQ calls.) And although we're not required to keep a log nowadays, an accurate station log is useful not only today, but tomorrow, too.

We all have things we like to keep track of: states and countries worked and confirmed; information for awards; or the names and addresses of our on-the-air friends. A well-kept station log is invaluable in your quest for the Worked All States or the DX Century Club (DXCC) awards. In addition to keeping a running list of states and countries, your logbook is the perfect place to keep detailed information on a wide range of subjects.

Your logbook is also a good place to keep notes on modifications to your equipment. Not only will the information be easy to find for future reference, it will be easier to note the effects of such changes by referencing contacts before and after.

How does your new loop antenna compare with your old trap vertical? Check out the signal reports in your logbook and you'll have a good idea!

DXers often refer to their logs when trying to work into specific parts of the world. When is the best time to work Africa in the winter? A quick check of last year's log entries will probably turn up the required information.

While you're at it, why not keep other changes in your log? When you upgrade, note it in your log. When you get a new rig or put up that long-awaited tribander, write it down. Ten years down the road, your log entries will bring back a flood of memories!

Speaking of memories, poring through your old logbooks can be a lot of fun. You'll come across your first QSO and remember how nervous you were, or you'll come across rare DX stations you've worked, pileups you've busted or the first time you worked someone special. Reliving those events is almost as much fun as it was when it happened!

Computers have become quite popular in ham shacks across the country—especially those belonging to contesters and DXers. If you have a PC in your shack, you might want to consider keeping your station log on your computer. A number of suitable programs are available. Check the ads in *QST*. Logging programs may also be available through your local club or computer user's group. If you're into programming, consider writing your own logging software.

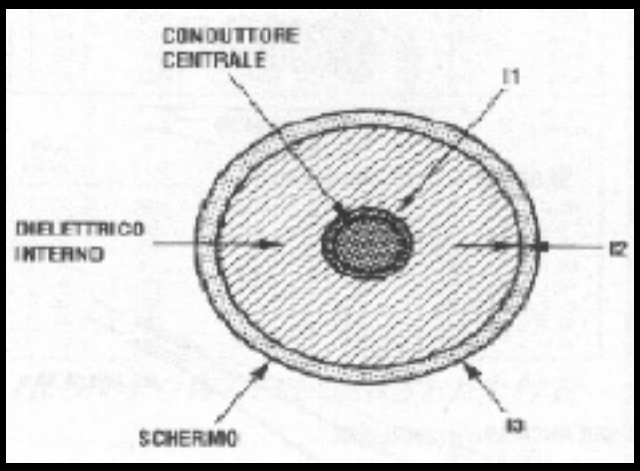
If you do go the computer route, remember to keep regular backups and a hard copy of your log information—otherwise, the benefits of having instant access may be lost if your data disks are lost or damaged.

If computerized logging isn't your thing, *The ARRL Logbook* is just what you've been looking for. Used by thousands (millions?) of hams over the years, the latest version is available from the ARRL for $3.50. It has room for nearly 1000 QSOs and includes useful information such as Q signals, a time-conversion chart, the ITU phonetic alphabet, an RST chart, international call sign prefixes and more.—*NTØZ*

GO TO HOME PAGE

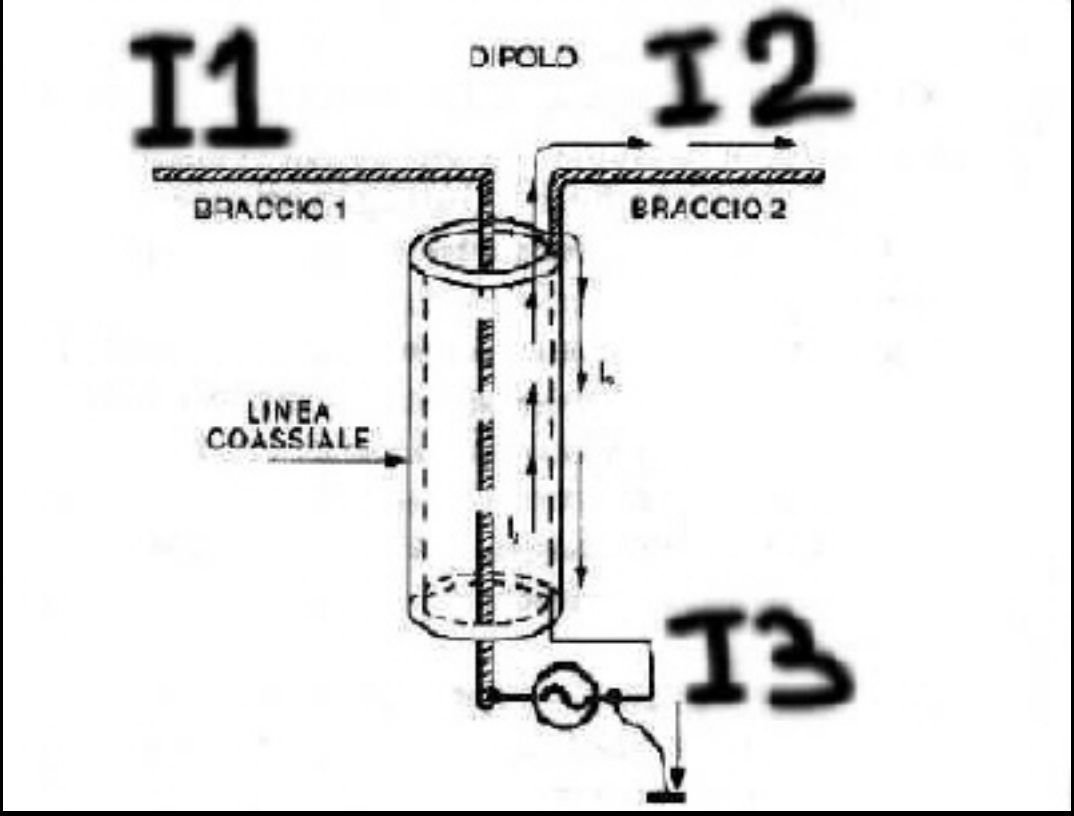
Teory Practice My Balun

ABOUT BALUN



May be baluns are still a mystery for hams; the only way to understand it is learning what it is and how to use it.
The word balun means balanced-unbalanced: it's used to adapt a balanced device to an unbalanced one; in a balanced device (as larger type of antennas) we have on both terminal the same voltage respect to the ground (if not so it's an unbalanced device);
a dipole with direct feed is balanced, a coaxial cable is unbalanced.
So, when we connect a balanced device to an unbalanced one the following occurs:

we have a dipole with coaxial cable direct fed (i.e. RG213); normally, transmitting, there are 2 currents on the cable:



1. **I1, which flows trough the central wire of coaxial cable and from transmitter goes up to the dipole;**

2. **I2, which flows (for the skin effect) on the inside part of the copper shield;**

The two currents, equal and opposite, humble itself and we have no radiation from coaxial cable.
The two currents cames on the dipole to be irradiated; part of it cames back; the one on the shield cames back trough the external side of the shield (no more the inside: so we have 2 current on the shield, I2 and I3); the value of this current (which we'll call I3) depends to the impedance value of the external side of coaxial cable respect to the ground (in a word if it will find high or low resistance);
if impedance will be high, I3 will find high resistance and its value will be low; if impedance will be low, resistance will be low and I3 value will be high; in this way I3 will radiate RF and the external side of the coaxial cable will radiate as a third wire of dipole: it's as we have a dipole with 3 wire (see picture 1); as consequence the radiation pattern will be distorced (see picture

2);
the big problem is that this new wire is often near TV and telephon cables irrading directly on those: so we have more probabilities to cause TVI.
This new wire, when its impedance is low, change the dipole impedance so we can have high S.W.R; this is the reason why (without balun) varying the coaxial cable length the S.W.R. changes.
Let's see now why use a balun; to delete this current I3, we need an high impedance for RF on the external side of coaxial cable; so we have high impedance casually for some cable length (odd multiple of 1/4 lamda) or using a balun: so the first reason to use a balun is TO AVOID THAT ON THE EXTERNAL SIDE OF COAXIAL CABLE FLOWS A CURRENT I3.
The simpler balun is a coil with some coaxial cable turns just belowe the antenna feed point: this inductance make the cable impedance (we mean the external side of cable shield) higher so that the RF current will find an high resistance and its value will be very low (that will not disturb I2 and I1 which flow inside the coaxial cable).

Continue...

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CHOKE BALUN

YOU ARE HERE: HOME > ANTENNAS > CHOKE BALUN

WHAT IS A CHOKE BALUN?

My friend Jim Duffy once said "Whatever antenna you chose, if it is fed with coaxial cable you should use a choke balun. This will prevent the feed line from becoming part of the antenna which can cause all sorts of problems. There are many designs to chose from. My favourite is an air core balun wound from coax".

Essentially, a choke balun is designed to "divorce" your antenna from the feed line. If your feed line is coaxial cable then you don't want it to be part of your antenna. You want to be able to deliver all your power to the radiator itself, i.e. "the antenna". A choke balun does this admirably.

Got a question on this topic?

If you are involved in electronics then consider joining our "electronics Questions and Answers" news group to ask your question

there as well as sharing your thorny questions and answers. Help out your colleagues!.

The absolute fastest way to get your question answered and yes, I **DO** read most posts.

This is a mutual help group with a very professional air about it. I've learn't things. Not for absolute newcomers to ask questions BUT it is an excellent learning resource for lurkers.

RELATED TOPICS ON AN Choke Balun

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Antennas

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[Loading Inductors] [Inductors and Loading Coil Current (Mobile and Loaded Antennas)]
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[End-fed vertical j-pole and horizontal zepp] [Ground resistance measurements]

Multiple Articles at:

Receiving: Contains information in many articles. Receiving antennas including Beverages and Beverage Antenna Construction, Loops, K9AY, Pennant, Flag, EWE, Slinky Beverage Antennas, vertical arrays, magnetic loop antenna, and terminated loop arrays

Transmitting: Contains information on various transmitting antennas

Single articles:

Balun and Core Selection: Contains information on core selection for transformers and baluns

Baluns, Sleeve balun: How they work

Baluns, transmitting: What they do, how they do it, and how to test them

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Crossfire Phasing: Contains information on cross fire phasing and why it is superior for broad bandwidth arrays

Combiner and Splitters: Contains information on Magic-T splitters and combiners, how they work and what they do

Detuning Towers: A quick explanation of how it works and the incorrect idea that you adjust for minimum current! Pass this along, it is a major error to tune for minimum current!

EH antenna: the E-H and CFA antennas. How they work.

Inductors, Loading: A brief tutorial on loading inductors

Mobile and Loaded Antennas: Small loaded antenna systems

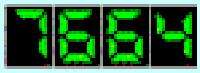
(Related page: Inductor spice model)

Omega and Gamma Matching: Contains technical information on Omega and Gamma matches and matching, impedance limits, component selection, component failures

Radiation and Fields: What the terms we use actually describe

Radiation Resistance: A revised (as of Feb 14, 03) tutorial on radiation resistance and how it is used and misused

Traps: Measurements and operation of traps and trap antennas. Including trap antenna loss.



since May 2004

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AMANDX PRESENTS

CHOOSING WIRE FOR AN ANTENNA

One of the most asked questions when it comes to antennas is what kind of wire should I use. The answer is as varied as the people who put them up. Over the last 28 years I have used or have seen used every kind of wire used to make an antenna. Just one disclaimer the information below is referring to antennas used for listening and NOT transmitting.

Stranded or Solid

I have seen both used with great success. Stranded wire is as a rule (and we all know rules are made to be broken) easier to work with and can at times be stronger. Stranded wire is usually more flexible so if you are not putting out a straight wire or bringing it into a house or apartment it may be superior to solid core wire. Solid wire is at times (especially with thinner wire) easier to break.

Coated or Bare

Wire with a coating is called insulated while bare wire in uninsulated. For antennas I do prefer the coated or insulated wire. The reason for this is that coated wire can be easier to work with as if it touches something conductive such as metal it is OK. Bare wire must be kept away from anything conductive to work properly. If bare wire comes into contact with anything conductive that material becomes part of the antenna. This can ruin an antenna's pattern or worse yet a ground system. This latter matter can cause an antenna to short out or worse yet cause damage to a radio due to static discharges. The choice would be yours as the coating or insulation will NOT decrease the signals received. If radio signals can go through your brick wall the tiny amount of plastic or rubber used to coat the wire will not bother your reception.

Thickness

The thickness of wire is measured by its gauge. The higher the gauge number the thinner the wire. I have used wire any where from 24 to 16 gauge over the years. Now the very thin high gauge 24 or 22 wire was used in pairs to give it some strength. No use putting out an antenna and have the first gust of wind tear it up. Wire of 18 or 16 gauge is quite good. I often use lamp cord or light indoor extension cord wire. This is insulated and cheap to buy at any hardware store. Hey you can be frugal and buy half the wire you need and split the wires to give you the length you need on a heavier gauge wire that is twinned like the lamp cord. If you live in areas that are subject to bad weather especially high winds or ice storms heavier gauge wire should be used so it can with stand the elements if the antenna is up in the air.

Copper vs Everything Else

In my humble opinion it matters not which you use. I have used everything from copper to aluminum to mystery metal over the years. I have never noticed any difference in the signal strength obtained from different types of wire. Go with the price on this topic depending on what you can obtain at your location.

Simple rules to follow:

Once you pick the wire you want make sure you have thought these points out:

If a wire is going to be left on the ground coated/insulated wire is a must.

If a wire is on snow you can use either as snow in an insulator.

If a wire is up in the air and away from anything conductive you can use either type.

If up in the air and near or touching anything conductive you must use coated/insulated wire.

If you live in areas that are subject to bad weather especially high winds or ice storms heavier gauge wire should be used so it can with stand the elements.

OK now go out and put up something. Experimenting is half the fun in this hobby.

Life span of an antenna...one man's joking opinion:

Well, generally it depends on hours spent listening.

The antenna converts electro-magnetic energy into electrical energy, which is basically electrons moving into your radio.
There are only so many electrons in each inch of copper wire, so when they've been sent downstream into your radio, the wire will become "ionized" and deteriorate and probably fall down. This explains why, when you come home one day, your antenna is on the ground (see below).

What happens to all those electrons, you ask. Well, they migrate into your radio and accumulate. In older tube radios, there was a "grid leak" resistor circuit which allowed the electrons to fall on the ground. Now you can't see them, but they're there. As more pile up, they slide into your back yard.

Tube radios, because of the "grid leak" last a lot longer than solid state radios, which stop working when enough electrons have piled up inside to short it out.

Now those electrons in your back yard want to get back into the copper wire, so they "pull" the antenna down to be re-united with it. Since the antenna is high, and they're on the ground, this attraction is not strong, but on a windy day, the electrons get lifted from the ground towards the antenna, pulling it down again. The wind oftens brings in free electrons from your neighbor's homes (from TVs, etc), so there may be a lot of these things around. If too many electrons get lifted up all at once, they overload the antenna, causing a heat mark, or worse getting back into the radio. Now this is why your antenna usually falls down on windy days.

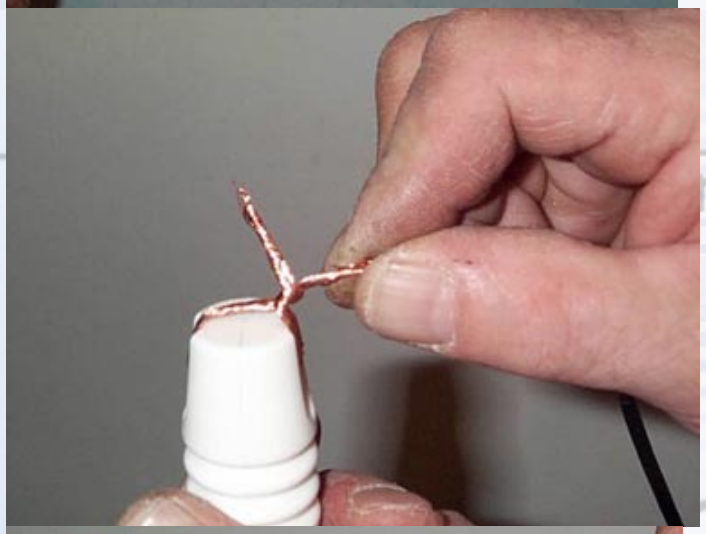
At least, that's how I understand it.

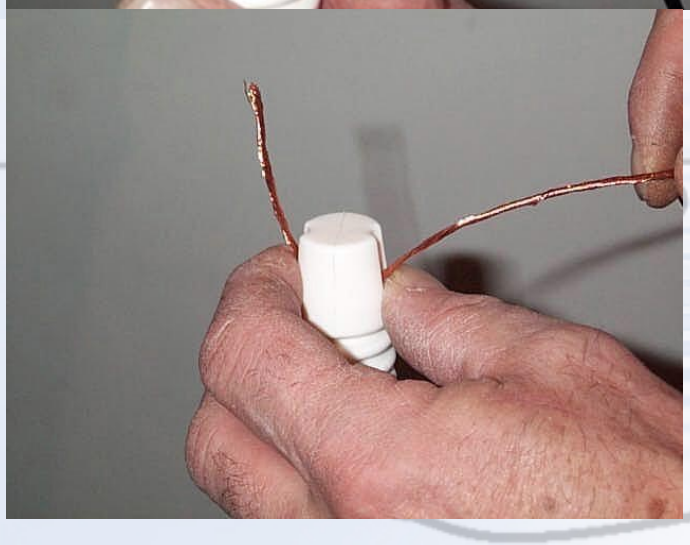
You can extend the life of your antenna by disconnecting it from your radio when you're not listening. But overall, 500 to 1000 hours spent listening will do in a longwire antenna.

It a joke OK!!!

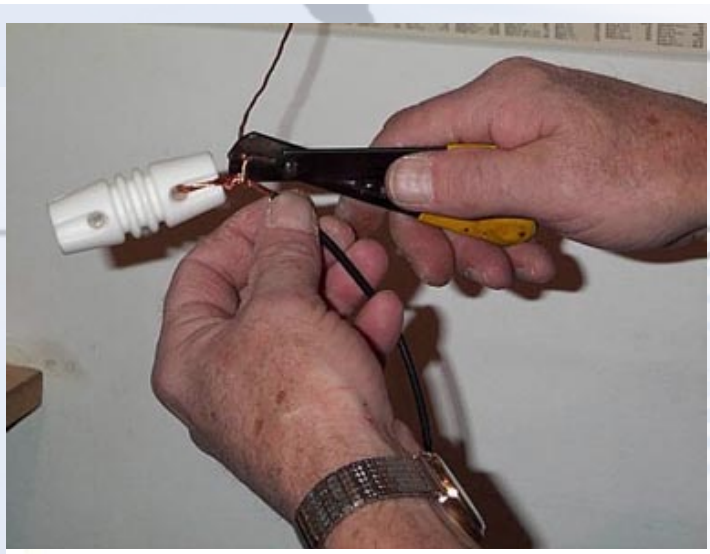
RETURN TO MAIN PAGE

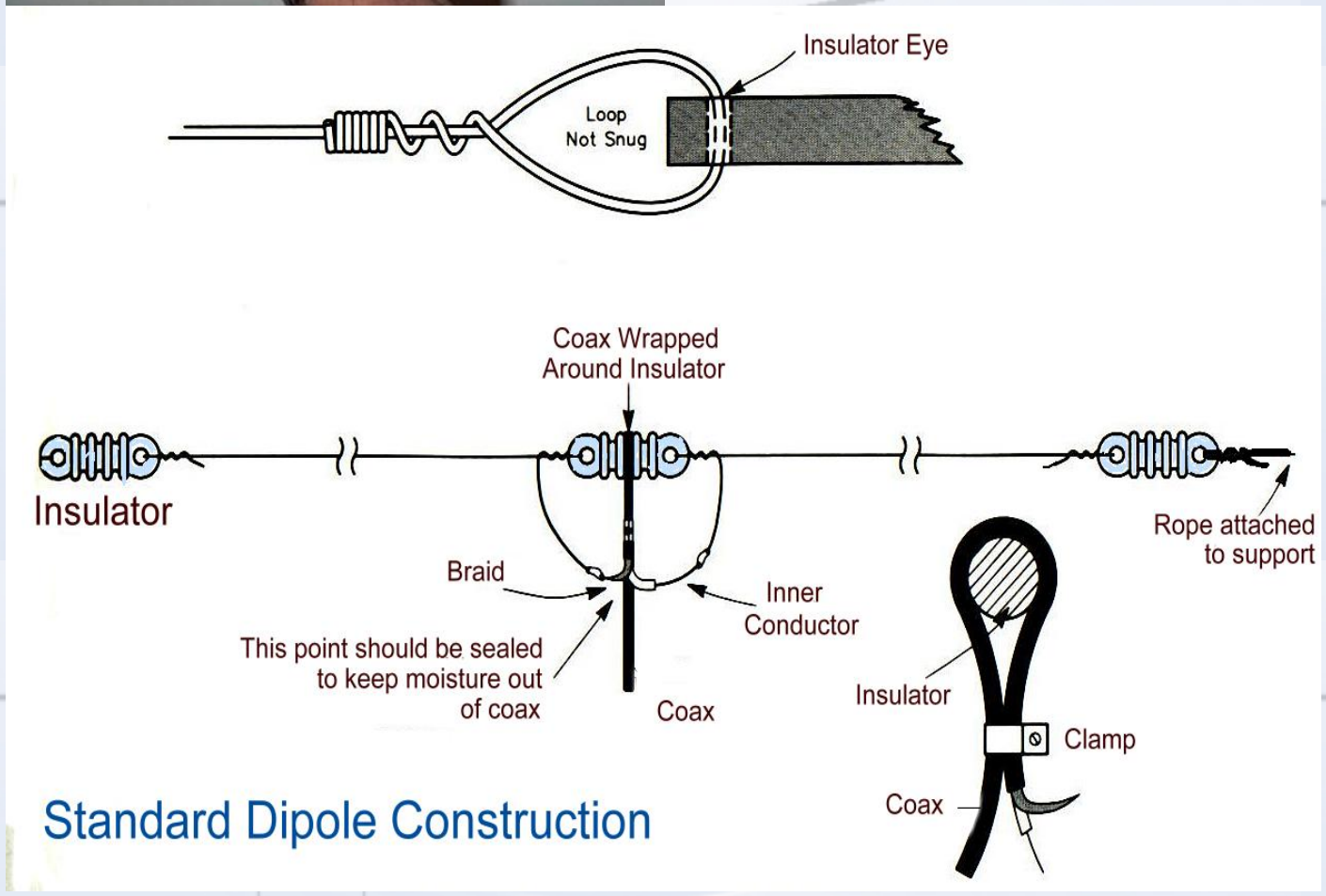
Remember On A Clear Day You Can Hear Forever





NorthWest Antennas



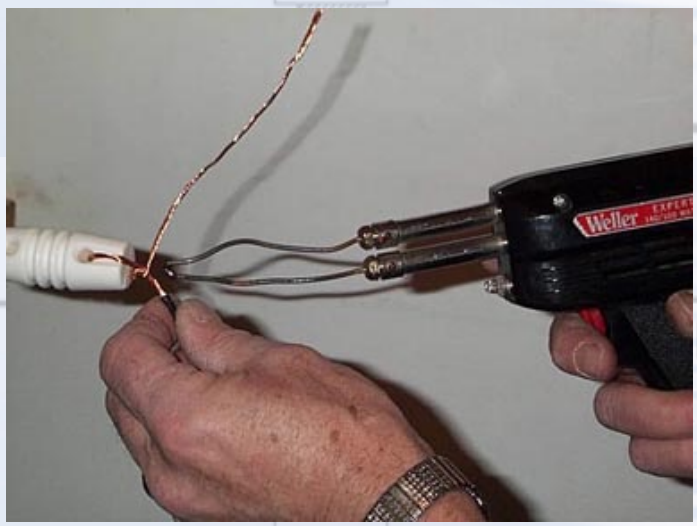


Installing your dipole end insulators.

Dipole links :

How high should by dipole be ?

I hope this information will help.
I wish you the best of luck.
73` Kevin



Total Length for Dipole Antennas

Frequency (MHz)
Total Length (Ft)

NorthWest Antennas Home Page

End insulators installing the fast and easy method here

Installing lugs/terminals on your coax

NorthWest Antennas eBay Auctions

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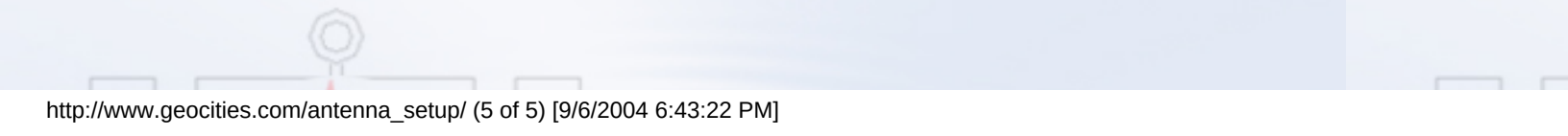


Tech Notes

Ground/Radial Systems

GROUND MOUNTING

A vertical antenna in its simplest form, is electrically equivalent to one-half of a dipole antenna stood on end. When the antenna is mounted close to the ground, the earth below takes the place of the "missing" half of the dipole. If ground conductivity is fair to good, a short metal stake or rod may provide a sufficiently good ground connection for resonant and low SWR operation on the bands for which the antenna is designed. This basic arrangement is shown in figure 1.

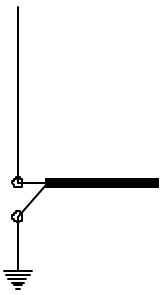


Figure 1

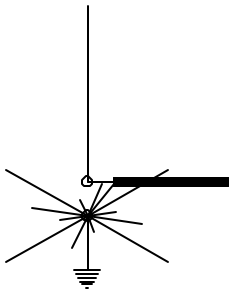


Figure 2

The way it works is that the capacitance between the vertical radiator and the ground causes *return* currents to flow along the earths surface back to the transmitter. If they have to come back along untreated lossy earth thy get back to the source greatly attenuated. This *return* loss is like a resistor in series with the antenna radiation resistance and will therefore affect the feed point impedance.

In almost every case the efficiency of a vertical antenna will be greater if radial wires are used to improve ground conductivity as in figure 2. It's important to note that there's no point in cutting radials to any particular length when ground mounting because the earth will detune them anyway. All you want to do is make the surface of the earth around the antenna more conductive than it is ordinarily.

If you can't copper-plate the backyard, the best approach is to run out as many radials as possible, each as long as possible around the antenna in all directions. Radials may be left on top of the ground however they should be buried for the sake of pedestrians and lawnmowers.

How long should radials be? A good rule is no shorter than the antenna is tall because 50% of your losses will occur in the first 1/4 8 out from the antenna. If you have more than a dozen radials, they must be longer to get the most out of them which is why the FCC specifies 113 wires each .4 8 for AM broadcast stations—the equivalent of a zero-loss ground plane. Obviously, for most ham work this would be overkill.

In some cases wire mesh (i.e. chicken wire) may be used as a substitute for radial wires and/or a ground connection, the mesh or screen acting as one plate of a capacitor to provide coupling to the earth beneath the antenna.

It should be noted that a ground rod is useful only as a d.c. ground or as a tie point for radials. It does little or nothing to reduce ground losses at r.f. regardless of how far it goes into the ground.

Bare wire, insulated, any gauge, it doesn't matter. The current coming back along any one wire won't amount to that much.

EFFICIENCY

The importance of reducing losses in the ground system can be seen from an examination of a vertical antenna's feedpoint impedance which at resonance consists of three components: antenna radiation resistance; conductor loss resistance; and earth loss resistance. An unloaded quarter-wave vertical antenna has a radiation resistance of about 35 ohms with negligible ohmic or conductor loss, but ground loss resistance may be very great if no measures are taken to reduce it, and in some cases ground loss R may even exceed the antenna radiation resistance. These three components may be added together to arrive at the feedpoint **impedance** of a **resonant** (no reactance) antenna. For the sake of illustration, assume that the ground loss beneath a quarter wavelength vertical antenna is 15 ohms, that conductor

loss resistance is zero, and that the radiation resistance is the textbook figure of 35 ohms. The feedpoint impedance would then be 15 + 0 + 35 = 50 ohms, and the antenna would be perfectly matched to a 50 ohm coaxial line. Since the radiation resistance is an index of the amount of applied power that is consumed as useful radiation rather than simply lost as heat in the earth or in the conductor, the radiation resistance must be kept as high as possible in relation to the total feedpoint impedance for maximum efficiency. Efficiency, expressed as a percentage, may be found by dividing the radiation resistance by the total feedpoint impedance of a resonant antenna, so under the conditions assumed above our vertical antenna would show an efficiency of 35/50 = 70%. As a vertical antenna is made progressively shorter than one-quarter wavelength the radiation resistance drops rapidly and conductor losses from the required loading inductors increase. A one-eighth wave inductively loaded vertical would have a radiation resistance of something like 15 ohms and coil losses (or trap losses for multiband antennas) would be in the range of 5 ohms. Assuming the same value of ground loss resistance (15 ohms), the feedpoint impedance would become 15 + 5 + 15 = 35 ohms and the efficiency would be 15/35 = 43%. From the above calculations it is clear that the shorter a vertical antenna must be the less efficient it also must be for a given ground loss resistance. Or to state the matter another way, more elaborate ground or radial systems must be used with shorter verticals for reasonable efficiency. If the ground loss of resistance of 15 ohms from the preceding example could be reduced to zero ohms, it is easy to show that the efficiency of our one-eighth wavelength loaded vertical would increase to 75%. Unfortunately, more than 100 radials each one-half wavelength long would be required for zero ground loss, so lower efficiencies with shorter radials must usually be accepted for the sake of convenience. In spite of their limitations, short vertical antennas over less than ideal ground systems are often more effective DX performers than horizontal dipoles which must be placed well above the earth (especially on the lower bands) to produce any significant radiation at the lower elevation angles. Verticals, on the other hand, are primarily low-angle radiators on all bands.

ABOVE GROUND (ELEVATED) INSTALLATIONS (rooftop, tower, mast. etc.)

The problem of ground loss resistance may be avoided to some extent by mounting a vertical antenna some distance above the earth over an artificial ground plane consisting of resonant (usually 1/4 8) radial wires. Four resonant radials are considered to provide a very low-loss ground plane system for vertical antennas at base heights of 1/2 8 or more. This arrangement contrasts favorably with the more than 100 radials for zero ohms loss resistance at ground level, and since 1/2 8 is only about thirty-five feet at 20 meters, very worthwhile improvement in vertical antenna performance can be realized, at least on the higher bands, with moderate pole or tower heights. At base heights below 1/2 8 more than four radials will be required to provide a ground plane of significantly greater conductivity than the lossy earth immediately below the antenna: even so, a slightly elevated vertical with relatively few radials may be more effective than a ground-level vertical operating over a larger number of radials if only because the former is apt to be more in the clear. Resonant radial lengths for any band may be calculated from the formula:

$$Length\ (ft) = \frac{240}{Frequency\ (MHz)}$$

Figure 3 shows the basic ground plane system for elevated verticals. Radials may slope downward as much as 45 degrees without any significant effect on operation or performance. Radials for different bands should be separated as much as possible and the far end of each radial insulated from supporting wires. Figure 4 shows a ground plane system that uses four resonant radials for 40 meters, another set of four for 20 meters, and a third set for 10 meters. A separate set for 15 meters is not ordinarily required because the 40 meter radials operate as resonant 3/4 8 radials on that band. At the lower heights the separate wires of this system may provide enough capacitance to ground to permit low SWR operation on 80/75 meters as well, but it is probable that at least one resonant radial will be required for low SWR on that band. It's important to note that cutting each conductor of rotator cable to a specific frequency will not work unless you separate it, angling each conductor away for most of its length because the longer ones will detune the shorter ones.

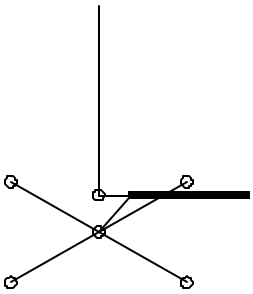


Figure 3

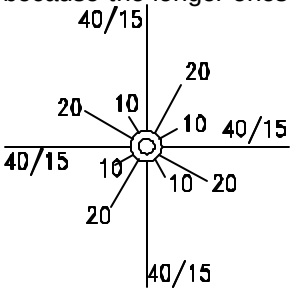


Figure 4

The 12-radial system of Figure 4 is a very good one, but it requires at least 12 tie-off points. Butternut has developed a multiband radial made of 300-ohm ribbon that resonates simultaneously on 40, 20, 15 and 10 meters. Four such radials offer essentially the same ground plane performance as the system of Figure 4 but require only 4 supports. These multiband radials plus additional wire for an 80 meter radial are available separately (our STR-II kit) or as part of the Butternut roof mounting kit (RMK-II).

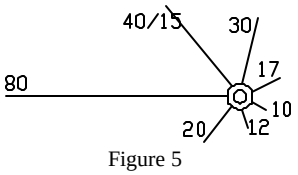


Figure 5

There are times when physical restrictions will dictate the use of fewer than four radials, and at least one

manufacturer recommends 2 radials per band, the radials for each band running 180 degrees away from each other. A simpler (and no doubt less effective) system is shown in Figure 5. Since only one resonant radial is used per band the antenna will radiate both vertically and horizontally polarized energy, and the pattern will not be completely omnidirectional. For true ground plane action and predominantly vertical polarization no fewer than three equally-spaced radials should be used.

Figure 6 illustrates the construction of a multi-band radial which is resonant on 40, 20, 15 and 10 meters. Good quality 300 ohm TV ribbon lead should be used (velocity factor is critical), and the conductors should employ at least one strand of steel wire to support the weight of the radial. Four such radials will be the practical equivalent of the system shown in figure four for operation on 40 through 10 meters.

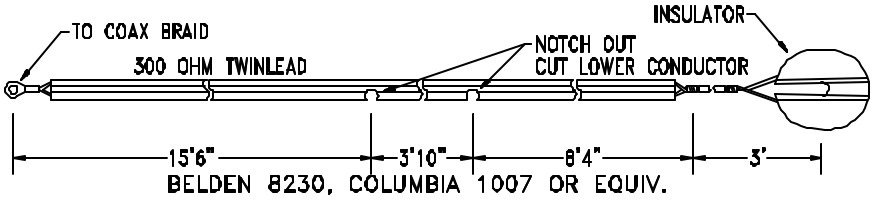


Figure 6

Regardless of the number of radials used in either elevated or ground level systems, all radials should be attached to the ground connection at the antenna feedpoint by the shortest possible leads. An elaborate radial system at ground level, for example, cannot be used with a vertical antenna on a rooftop or on a tall tower, for the length of the ground lead would effectively become part of the antenna, thus detuning the system on most or all bands.

METAL TOWERS AND MASTS

If a metal mast or tower is used to support a vertical antenna all radials should be connected to the mast or tower at the ground connection of the antenna feedline. This is because one of the functions of a resonant radial is to detune a supporting metal structure for antenna currents that might otherwise flow on the structure and thus turn the vertical antenna system into a vertical long wire with unwanted high-angle radiation.

OTHER MOUNTING SCHEMES

In cases where a resonant vertical antenna may neither be ground mounted nor used with an elevated ground plane, operation may still be possible if connection can be made to a large mass of metal that is directly connected or capacitively coupled to the ground, e.g., central air conditioning systems or structural steel frames of apartment buildings. Some amateurs have reported good results with vertical antennas extended horizontally or semi-vertically from metal terraces which serve as the ground connection. Alternatively, a quarterwave vertical may be window mounted if a short ground lead to a cold water pipe or radiator can be used. If a long lead must be used, tuned radials may be required for resonance on one or more bands. Great care should be exercised in such installations to avoid power lines and to keep the antenna from falling onto persons or property.

MOBILE HOME AND RV INSTALLATION

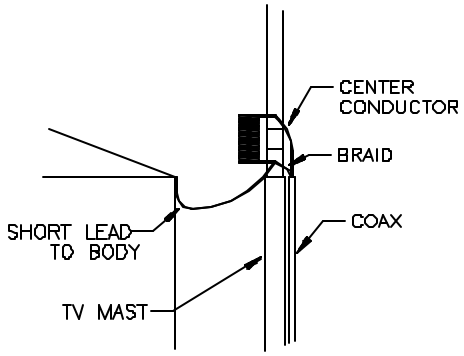
The principles of vertical antenna installations for use on mobile homes or RV's are the same as for other installations, and they all boil down to two main considerations. The first is that of erecting the vertical in the clearest possible spot, away from obstacles (including the MH or RV) that can interfere with radiation from the antenna. The second is that of installing the beat possible ground system beneath the antenna in order to minimize losses from r.f. currents flowing in the earth below the antenna. Fortunately, the metal bodies of both MHz and RVs can be used as highly conducting ground planes for vertical antennas in exactly the same way that automobile bodies, etc., provide the ground system for shorter vertical antennas for mobile operation. The metal body of an automobile, MH or RV may be viewed as one plate of a capacitor. Since the surface area of even a small automobile is quite large and in close proximity to the earth, its body is tightly coupled to the earth below and may be considered simply as an extension of the earth itself—a kind of hill as far as radio frequencies are concerned, but one having higher conductivity than the earth itself. RVs and especially MHz, having much greater surface area, will therefore provide a more extensive and effective ground system than a large number of radial wires occupying the same space as the MH or RV.

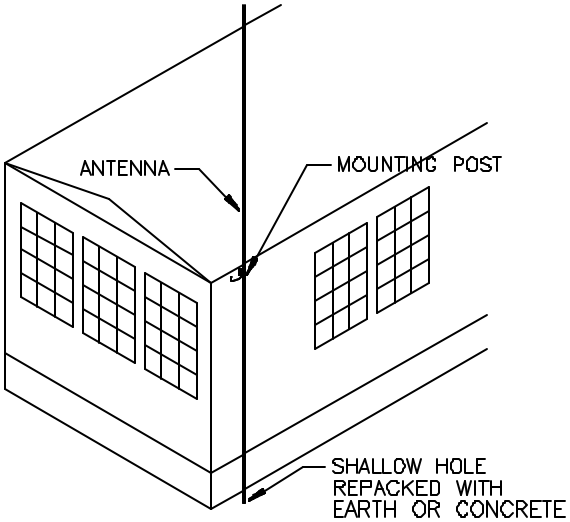
As in mobile installations, a vertical antenna may be mounted almost anywhere on the body of the vehicle or MH and made to operate with reasonably low VSWR, but it is generally considered that the best possible location for a mobile antenna is in the middle of the roof of the vehicle, i.e., at the center of the vehicle's ground plane and at a point where the antenna will not be in the "shadow" of any part of the vehicle. It is not usually convenient, or even practical to install a relatively tall vertical on the roof of an RV or MH for any number of reasons, so the next best procedure would be to install a vertical antenna with its base at the same level as the roof, preferably near the middle of one of the longer sides.

The exact way in which this may be done is a matter of convenience, but a short mast extending from ground level to the roof of the MH and RV and placed alongside the building or RV would provide a stable and sturdy support with a minimum of mounting brackets and other modifications to the RV or MH. For portable operation such a mast could simply be lashed alongside the RV with the base in a shallow hole in the ground for additional support, and there would be no harm in extending the mast a few inches above the roof level to permit attachment of ropes which could be used to hold the mast firmly against the side of the vehicle and to prevent side sway.

This system has been used successfully with various types of RVs, travel trailers and even passenger automobiles during portable operation. For "L" shaped mobile homes a vertical antenna should be placed in the corner of the "L" so that the metal roof will provide groundplane coverage over 270 degrees.

In all cases the base of the vertical antenna should not be more than a few inches away from the MH or RV so that the shortest possible lead may be run from the ground connection of the antenna to the metal body, as the length of this ground lead will effectively lengthen the antenna itself on all bands, and detuning can occur in some cases. A good electrical connection between the body of the RV or MH and the antenna is important, and in the case of mobile homes it would be a good idea to make sure that good electrical contact exists between the different parts of the metal body. Discontinuities can often lead to the production of harmonic radiation and TVI. The essential circuit connections are shown in the diagram above.





For permanent installations the bottom of the mast may be set deeper in the ground, and concrete may be used for greater strength and stability. The upper portion of the mast should be securely attached to the side of the building. Steel TV mast sections are readily available in lengths of ten feet and the mounting posts of Butternut HF verticals will slide into those which have an outside diameter of 1 1/4 inches and a wall thickness of .058 inches. Other vertical antennas may use different mounting techniques and requirements, so be sure to select a mast that will be suited to the particular situation. The main point to keep in mind is that the mast should not extend more than a few inches above the level of the roof so that the ground lead may be kept short.

LIGHTNING PROTECTION

Modern solid state amateur equipment is particularly vulnerable to damage from lightning or static induced transients that may appear on transmission lines, and conventional air-gap lightning protectors may provide no real protection at all for solid state gear. A line of very effective lightning and static protectors has been developed by ALPHA DELTA COMMUNICATIONS, P.O. Box 571, Centerville, Ohio 45459, for use with solid state equipment, and since these devices feature much faster transient discharge times than earlier designs, they should be investigated for possible use with all vertical and other antenna systems.

Inductors and Loading Coil Current (Mobile and Loaded Antennas)

[Home] [Up] [Independent Measurements] [Inductor operation]

Related pages:

Mobile antenna FS comparisons Loading Inductors Inductor Spice Model

W7EL's Measurements and comments Constructing an RF Current Meter

Much of the data below also applies to inductors in equipment, such as tank circuit.

If you arrived here from a link from K3BU, *welcome to W8JI.com I hope you enjoy your visit.* ***If you are not linked here from loading coil current articles on another web site, please skip the blue-text immediately below. It is meaningless.***

If you arrived here from a link from K3BU, I want to apologize for the personal nature of that page and this response. Why is this response necessary? None of us like to see false or out of context claims or personal attack articles. Many of the comments from below were removed from context and quoted to mean something else in another internet article. It is important to read everything ***IN CONTEXT****.*

How Does an Inductor or Loading Coil Work?

The most basic answers are:

What does the coil do? A loading coil does not replace a missing fraction of a wavelength. The coil simply inserts a series inductive reactance that cancels capacitive antenna reactance. When a 150 ohm reactance inductor is inserted in series with a 150 ohm capacitive load (like an antenna), only the resistive parts remain.

What determines current distribution in a loading coil? The capacitance to the outside world and the impedance above the loading coil. The current in any inductor would be equal at each end except for displacement currents, which are "imaginary currents" that flow through capacitance.

How much difference is there in loading coil current entering the coil and loading coil current exiting the far end? If the antenna beyond the coil has a low self-impedance compared to the impedance of the shunting capacitance from the coil to "ground", the currents at each end of the coil will be essentially equal. It has NOTHING to do with electrical degrees the coil compensates for. In other words if the portion of antenna above or beyond the loading coil is long or has a large area compared to the physical size of the coil, current is essentially equal throughout the coil.

What does significant current taper in the loading coil indicate? It generally indicates a poor antenna design, where the loading coil (or any other application using an inductor) has high stray capacitance to other areas of the antenna system (like the groundplane) compared to the capacitive reactance of the antenna beyond the inductor. Significant current taper indicates a poor loading coil or poor antenna design.

click to view typical installation measurements

The Difference Between a Loading Coil and a Normal Inductor

There really isn't any difference, except the location where the coil is used. Mobile loading coils or loading coils for short antennas often have very high reactance. They have small amounts of capacitance at the end, and so stray capacitances are more of a concern. Stray capacitance from turn-to-

turn increases circulating currents and has the effect of increasing inductance and effective resistance at the expense of reducing bandwidth and Q. Stray capacitance to the outside world causes the coil to behave like an L-network, and transform impedances instead of providing a series reactance. This is why the optimum form factor of a coil becomes longer compared to the diameter with any inductor having very high reactance.

Inductors with low reactance are less critical of stray capacitances. Optimum form-factor in a low reactance inductor leans towards a short coil with the diameter nearly equaling length. In tank circuits or loading coils used with longer antennas or with capacitance hats, an optimum inductor is shorter and larger in diameter.

Most optimized inductors fall between 1:1 and 4:1 length to diameter ratios, the exact value depending on the terminating reactances. A coil is a coil, they behave the same way regardless if used as a loading coil or a tank inductor.

Common myths about inductor behavior:

One common myth is loading coil current is reduced as it passes through the coil. There are two reasons cited for this. One idea is the current is reduced because the loading coil replaces a certain amount of "electrical degrees" of antenna area, like the current taper in an antenna. The other idea is that series loss resistance causes a current reduction.

We often find inexperienced builders of 5/8th wl antennas think the "loading coil" needs to contain 1/8th wavelength of wire in order to make a "3/4wl antenna" and thus cause a low feed impedance. In other cases, some claim a half-wave of wire wound on a form causes a 180-degree delay, and is useful for phasing in a collinear array.

The basic flaw is the above ideas do not account for what happens in a coil with mutual coupling between turns. The flawed viewpoint is that current goes in one end, winds its way around through the physical length of wire in the coil, and after a time delay comes out the other end. There is a physical mechanism that prevents this, as we might intuitively think, from actually happening . The mechanism is the magnetic field in the coil!

What Really Happens

When current flows in the transmitter-end of the coil, a magnetic field is created. This magnetic field causes charges in the other turns to instantly move. This effect ripples through the coil at light-speed, just over 186,000 miles per second. As long as the magnetic flux coupling is high, the delay through the coil is the speed of light over the physical length of the coil. The electrical delay is the physical length of expressed the coil expressed in degrees at the operating frequency.

(Another interesting effect occurs. The increasing magnetic field sets up an "opposing voltage" as it cuts across conductors. This opposing voltage, created as the field expands, is what causes the current to rise slower than the applied voltage. If the exciting voltage is decreased the field collapses, and now the voltage changes polarity and aids current flow! If we don't allow the current to flow, the voltage will rise until it does. This is what causes the kick in a relay coil when we open the relay coil path, or the spark in an ignition when the points abruptly open.)

In an RF system, the physical size of the coil actually does add some "antenna effect". For example, on 160 meters the wavelength is about 550 feet. 1.5 feet is about one electrical degree. A skinny one foot tall coil, with negligible stray capacitance, would have about 0.67 electrical degrees phase delay. This delay occurs because to coil occupies a physical length of .67 degrees. Current at each end would be almost perfectly equal, the taper would be about what we would expect for a fractional-degree-long coil.

(*In the real world, all components have some stray capacitance and flux leakage, so they have a different amount of electrical length and current taper than the "negligible capacitance" case. In good coil designs, the capacitance and leakage is small and can be ignored. I'll show you measurements later to prove this.)*

Now let's look at an extreme case. If the entire antenna is "coil", like a helically wound antenna with no top hat or stinger, current would be reduced to nearly zero at the open end. This is because distributed capacitance over the length of the antenna is fairly high, the shunting capacitance has a low impedance compared to impedance at the end of the antenna, and current is diverted to ground in the form of displacement currents.

Compact loading coils are another matter. In many cases phase delay is negligible or immeasurable by normal methods...flux coupling is nearly perfect. A good example would be a relatively compact toroid or a compact nearly-square L/D ratio loading inductor. I've found it impossible to measure the current taper in a toroid and very difficult to measure in a compact air-core loading coil. (The opposite extreme would be be a perfectly straight wire with no folds or bends or the helical antenna described above.)

In the case of the toroid or compact coil, the behavior would be such that doubling the turns nearly perfectly quadruples inductance. If we doubled turns and inductance simply doubled or increased at a much faster rate, we should know the coil is in a mode other than a pure inductor mode. This is a ***strong*** indicator inductor operating Q is less than optimum, and the inductor might behave less than ideally in critical applications.

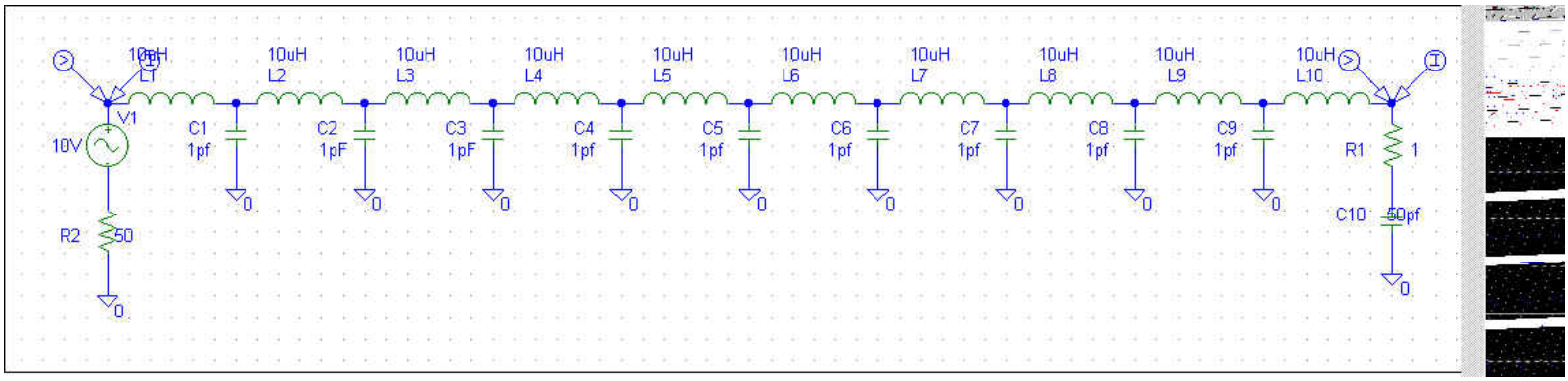
As a matter of fact, observing inductance change while adding turns can be an excellent test for flaws or shortfalls in system design. A linear increase in inductance when adding conductor length indicates design problems.

A perfect impedance squaring effect indicates minimal electrical phase-delay, or "antenna length" of an inductor. Impedance squaring as turns are doubled indicates the undesired inductor stray capacitance has a high reactance compared to the antenna system beyond the loading coil. Of course there can be exceptions, but it is a good general rule that large current taper indicates the loading system is much less efficient than necessary.

Making a Delay Line

It's certainly possible to make a delay line from a coil without opposing flux, but doing so requires stray capacitive reactance to be significant compared to the value of distributed inductance in the coil. This would occur in a very long helice, a very large diameter helice or loop, or an inductor near a large metal counterpoise or ground plane.

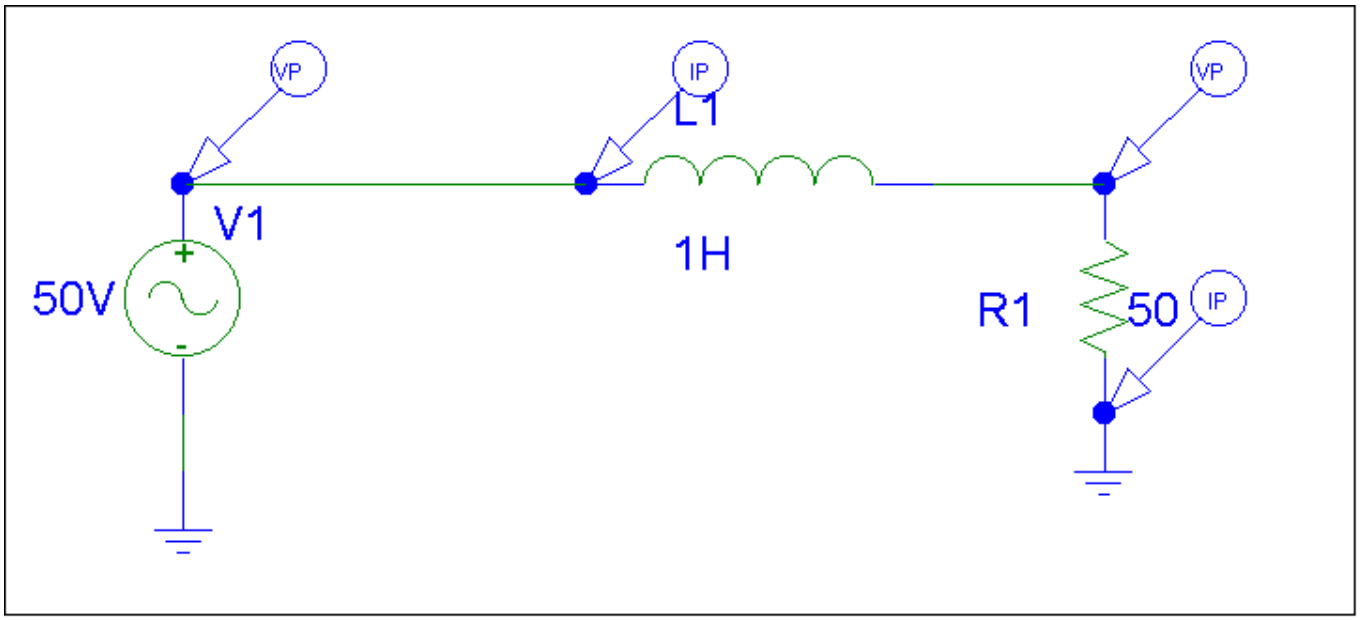
It's important to remember unless a coil is "stretched out" or "expanded" a great deal, the phase delay will not even be close to the physical conductor length. (The exception could be if you had so much capacitance the inductor acted like a series connected string of L/C/L networks as shown below).



In any case while this effect might be good in a collinear antenna or plate choke (assuming you do it right) it is a BAD effect in a short loaded antenna!

Inductor E/I Phase shift

An inductor delays the flow of current in relationship to applied voltages as the magnetic field inside the coil expands. Voltage increases before current starts to flow. This phase relationship between voltage and current is often confused with time-delay phase in the inductor. Say we have this simple circuit:



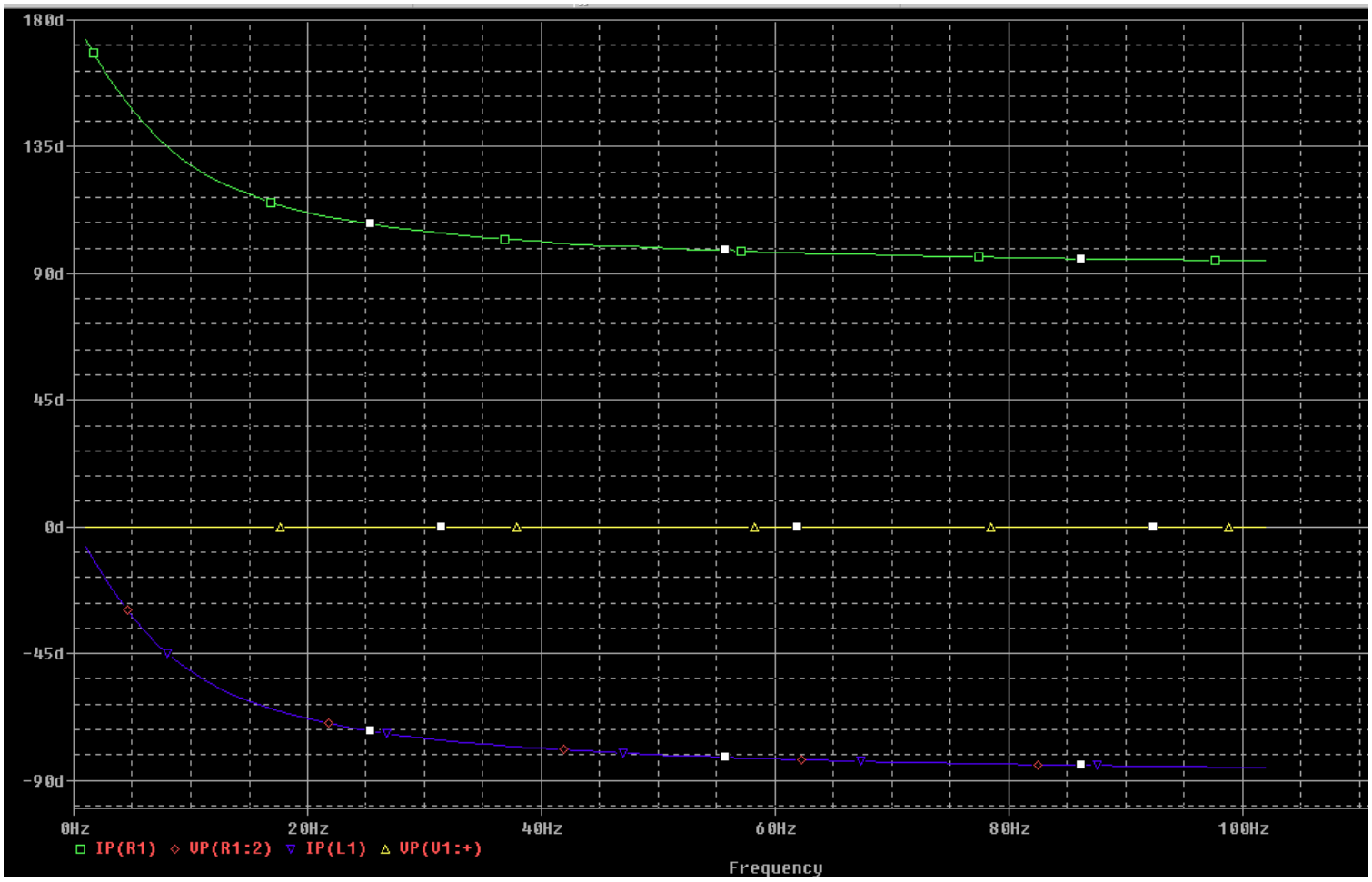
Current and voltage at V1 will be out-of-phase by the effect of L1 "charging" with magnetic flux. Current appears AFTER voltage rises, and falls after voltage falls. Current in R1, however, is exactly in phase with voltage across R1. That's because the voltage across R1 is always E=I*R.

Every component must follow the Laws or electrical rules established for that component.

The current in R1 is delayed from VOLTAGE rise in V1 by the voltage to current phase delay of L1. This does result in a time delay in relationship to voltage rise at V1, but there is NO current time delay through L1! V1, L1, and R1 all have the same peak current at the same time!!!

The notion that antenna loading coil delay current by the same time as they delay response to increased voltage is obviously nonsense.

Here is a graph of phase delays in the above system:



Current in the load, generator, and inductor all exactly track in the same relationship from dc up. There is no "phase delay". The generator voltage is a straight line different than current, and this indicates the generator sees a "reactive load".

The Misplaced Notion

Proponents of the idea that coils replace "antenna length" so far have been unable to define a set of rules or logical reasons why a current reduction and electrical-degree phase delay related to antenna area "replaced" would occur in a two-terminal component. While a long inductor with poor flux linkage from end-to-end or an inductor with low values of stray capacitive reactance to a groundplane compared to series impedance can cause SOME current inequalities or phase delays, the amount is normally immeasurable with normal thermal current meters with a reasonable coil form factor and termination above the coil. The amount of current taper actually rivals the disturbance of the system by adding the measurement device, unless we are very careful in how we construct the measurement device.

Both W7EL and myself made independent measurements that show the fact the coil or inductor is in an antenna has nothing to do with current distribution in the coil. Some people have actually incorrectly reported W7EL's data! Here's what he had to post on rec.radio.amateur.antenna to correct misrepresented claims.

The Need for a Measurement

An article on E-ham claimed measurements proved a new concept about loading coil current. The E-ham article put forth an idea that current disappears as it moves through a loading inductor without a mechanism like displacement currents providing a path. This claim conflicts with established component behavior, so it would indeed be fascinating if it were true! One of the claims supporting the idea that coils in antenna work differently than coils in circuits was that a non-radiating toroid loading inductor showed a current taper when used in an antenna.

I recently constructed a calibrated current meter that slips over whip antennas and masts, and is for all practical purposes totally immune to variations in voltage in the system. It also is mostly plastic, and has minimal effect on stray capacitance of the antenna. The resonant frequency and currents are not significantly perturbed by measurements with this meter. When I added a similar meter used in the other tests, resonant frequency shifted significantly! This is a sure sign the meter's capacitance or inductance is affecting the system.

In late December 2003 and early January 2004, I made additional measurements of loading inductor currents. The results clearly agree with the analysis that had been presented on this page since early 2003. Without displacement currents, currents into and out of a loading inductor *are equal.* That is a hard rule, it agrees with theories defined by people much smarter than me, and I believe it is unbendable unless the works of Faraday, Maxwell, Ohm, and Kirchoff were incorrect.

A sample of measurements above and below the loading coil with various antenna above the coil (current as percent of reference) follow:

	Toroid with hat	Small 2x2" coil with 24" hat up 24"	Long 12x3" coil with 24" hat up 24"	Long 12x3"coil with 6' whip
Current below	100%	100	100	100
Current above	100	94.4	73	76%

	Toroid with whip	long 12x1.5" coil with 6' whip	long 12x1.5" coil with 24" hat	Small 2x2" coil with 6' whip
Current below	100%	100	100	100
Current above coil	100%	79%	75%	96%
Current in whip 1ft above top of coil		73%		

The most revealing thing was how noticeably small changes in stray capacitance near the middle and top of the loading coil affect current distribution. It was quite evident hanging a large meter on each end of the coil would greatly perturb the system.

Clearly we do NOT want:

- **A large hat just above a large coil**
- **A long large coil and a short whip**
- **A coil near large sheetmetal**

More data along with photographs will appear on a new page over the next month or so. Until then, I can assure everyone the conventional theories presented below are accurate, and the theory that "electrical length" the coil "replaces" is incorrect. Loading coils indeed behave like any other inductor in the world.

Independent measurements by a reliable engineer have agreed with my measurements above. Anyone doubting my data need only read the following e-mail from W7EL.

On to the old text in this page:

The Incorrect Assumption

Another commonly misconception is, since voltage increases at the far end of the loading coil, current must logically decrease. After all, we have a fixed amount of power and voltage has increased. The assumption is:

1.) We multiply voltage times current to get power.

2.) If voltage increases current must decrease.

Unfortunately, this is ***not*** correct in reactive systems! Simple P (power) = I (current) times E (voltage) only works when the system is non-reactive. This condition only occurs at resonance, and only ***below*** the loading coil at the antenna feedpoint!

In a reactive system, like in a mobile whip above a loading coil, voltage and current are no longer in phase. As a matter of fact, voltage and current can closely approach being 90 degrees out-of-phase when the whip is electrically very short. Since the antenna area above the loading coil is highly reactive (voltage is not in phase with current), we can not multiply voltage times current without considering phase differences.

You may have heard the term "reactive power" or VAR (volt-amperes-reactive). Reactive power is voltage times current without consideration of phase angle. We can have kilowatts of VAR power with only a low power transmitter, and that is what we actually have in the reactive part of the small antenna.

Coil Q and Changes in Efficiency

Current taper or reduction has been cited as a reason coil "Q" has little effect on signal level in mobile systems. Speculation is only the first few turns of the loading coil carry significant current because the coil "leaks" magnetic fields and radiates, and this is why the coil Q has little effect.

Another idea proposes the loading coil "makes up" a certain missing part of the antenna. It goes on to conclude the loading coil can be accounted for in "electrical degrees", making up the "missing difference" in antenna degrees. This isn't true either. The inductor doesn't know where it is and suddenly change from "x" ohms reactance to electrical degrees! It responds to AC currents and voltages as any inductor in any circuit does. It doesn't suddenly change measurement units.

As an example of this, try to define a 45-degree electrical length inductor at 1.8 MHz. That would mean it is a capacitor at 3.61MHz, where it is over 90-degrees long! How many turns at what length and diameter is a 45-degree inductor?? Where is a formula that allows converting a given size inductor to electrical degrees? This shows how useless and meaningless that definition is!

The inductor adds a certain amount of series reactance, that's all. A 300uH inductor is not 20-degrees long, nor is it 80-degrees long, so far as radiation goes unless it is really that long ***physically***. It is a certain number of ohms reactance at a certain frequency, or a certain number of units called Henries. It is not "electrical degrees" that it adds, it is a non-dissipative reactance (in combination with a loss resistance because of finite quality) at a certain frequency!

A loading inductor can "insert" a large amount of phase shift, but the phase shift is between voltage and current. The only exception to this would be if the inductor had considerable distributed shunt capacitance to the outside world, and acted like a string of series inductors (with the antenna) and shunt capacitors (shunting to the ground system). In that case we could expect coil Q to be extremely low, since it would be the electrical equivalent of a lossy transmission line. That's either an awful loading coil, or it is a less-efficient helical loaded antenna!

The Correct View

Another group of people don't argue against established and proven circuit theory. They understand charges flowing into one end of the loading coil must have someplace to branch off (a virtual third terminal), or they must flow out the other end. Without that additional "virtual" path, charges flowing into the coil would always equal charges flowing out. This is true regardless of radiation, losses, or induction fields.

This makes perfect sense when we think of any dc circuit, antenna, or RF system. Electrical rules are satisfied, the system behaves as it does in the real world.

There is very little change in current, unless the coil is physically very long compared to the rest of the antenna above the coil or unless the coil is laid right against "grounded" conductors and the whip above the coil is very short. This fits perfectly with helical verticals, where the coil is "stretched out" over the length of the antenna.

It also agrees with base loaded antennas, which have nearly as much current into the antenna above the coil as at the feedpoint. It agrees with center loaded antennas, where current below the coil is essentially uniform and the whip above has triangular distribution.

Current ***can*** be different in various areas of an inductor, but ***only*** if shunting capacitances (impedances) to the outside world are significant compared to load capacitance (impedance). Another condition where current can vary substantially is with operation near the condition of self-resonance in what is normally considered or defined as a "series-resonant" mode. This would be a very poor and inefficient loading inductor, such as when a 160-meter antenna is used at a secondary resonant frequency in upper HF.

Circuit Model of a Mobile Antenna

The above model shows what might be a typical mobile antenna installation.

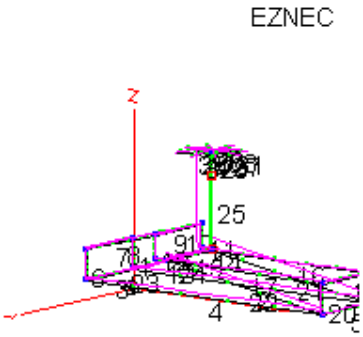
1. Rgnd ground resistance of vehicle normalized to feedpoint
2. V1 coaxial feedline
3. C2 base capacitance
4. Rrbase Radiation resistance of the base area of the antenna
5. L1 loading coil
6. Rcoil coil equivalent series loss resistance
7. C3 coil shunt C to ground
8. Rr top area radiation resistance
9. Cant equivalent antenna capacitance above coil area

My 160 Mobile Antenna

I've worked all continents except Africa while 160-meter mobile. I have CW contacts at over 10,000 miles, and SSB as far as Europe (4000 miles). My mobile antenna consists of an eight- foot antenna with a six-foot hat (made from surplus Ford or Chrysler car antennas) at the top. The loading coil is at 5 feet.

This antenna has been on the truck for thousands of miles, without mechanical failure. It is mounted at the upper left corner of the truck bed, about one foot back from the cab.

The following is a model of the current antenna system on my Ford F-250 HD long bed super cab truck:



The base impedance in the model is:

Frequency = 1.854 MHz.
Source 1 Voltage = 24 V. at 2.35 deg.
Current = 1 A. at 0.0 deg.
Impedance = 23.98 + J 0.9853 ohms
Power = 23.98 watts
SWR (50 ohm system) = 2.087

Actual measurement at my Johnstonville, GA farm in open flat pastures on August 17 at 8AM. Wet soil 25.8 ohms 0j base impedance, pretty close agreement to EZNEC model and earlier data! (I did have to adjust the model for very low ground conductivity, otherwise the resistance was far too low. It's my belief that NEC-2 underestimates ground losses in small radial or counterpoise systems that are close to earth.) Earlier text shows a base resistance of ***28 ohms***, that was dry soil with a slightly different loading coil and antenna.

The modeled current distribution for 1-ampere applied at the base (in 1-foot intervals) is:

1ft= 1.0031

2 ft= 1.0091

3ft= 1.0178

4ft= 1.0318

<Coil>

5ft= 1.0175

6ft= .97512

7ft= .92984

8ft = .89522

Measuring the current into and out of the loading coil with a small thermocouple RF meter, I detect no difference This is in close agreement with the model.

The efficiency of this antenna knowing coil Q, radiation resistance, and base resistance calculates just under 1 percent. The model indicates about 1/3 percent efficiency. This is reasonably close.

Removing the hat (in the model only) shows the following changes:

1ft = 1.0043
2ft = 1.0133
3 ft= 1.0279
4 ft= 1.0566

<coil>

5 ft= .95508
6 ft= .72232
7 ft= .27813
8 ft = open

I haven't tested the above, but with the same loading coil loss resistance the model says efficiency is now around 3dB worse. Removing the hat, with NO change in coil resistance, shows nearly loss nearly doubles. Of course the coil resistance would increase, because the loading coil nearly quadruples in size. Bandwidth is less and efficiency is less, even if I could maintain the same coil resistance.

Examples of Unequal Current

In the above models, we see that current into and out of the one-foot long coil is about the same. There is only about ***2% change in current*** even though the coil occupies 12% of antenna length in the "hat-loaded" antenna, but in fairness I couldn't resolve that change with a reasonably good RF current meter.

The model predicts ***10% change in a non-hat antenna***, but I never measured that antenna to confirm it.

Clearly there is no basis to the claim current is high only in the first few turns of an inductor, or that current tapers in relationship to "electrical degrees". The most accurate way to state the effect would be to say: "When the loading coil is short and the capacitance of the antenna beyond the coil is reasonable (in this case 3000 ohms Xc or less), there is an immeasurable reduction in current in the coil. When the required loading reactance is very high (in this case 8000 ohms), the reduction in current is about what we would expect for an equivalent length of antenna replacing the coil."

Degrees Vs Radiation Resistance

This upper four feet of this antenna resonates near 24 MHz with the hat. We can assume it is 90 degrees long at 24 MHz, which would translate to 6.9 degrees on 1.85 MHz. Following that same logic, this would mean the loading coil would be about 83 degrees long electrically. Using the incorrect logic proposed by others where the loading coil "makes up the difference in electrical degrees", there would be almost no current past the loading coil. Obviously this is not the case, the loading coil has very little "electrical length". As a matter of fact, the electrical length is about equivalent to the physical length!

This goes back to radiation theory, and my favorite saying: "Five hundred feet of wire in a one foot long tube is still one foot of antenna". Some CB manufactures sell antennas to consumers with the claim they use 5/8 or 3/4 wavelength of wire in an eight-foot fiberglass whip, so the antenna has more gain. Obviously this is not true. Let's not let such silly claims spread into amateur radio!

Related topics:

Inductors

The spice inductor model shows one example of how unequal current is created. The model demonstrates a coil having significant distributed capacitance to the point of current return in the system compared to terminating impedance of the coil. In a monopole this return path would be to the groundplane, or anything closer to the potential of the groundplane than the area above the loading coil's position in the antenna system.

Another Practical Antenna Example

Let's assume we have a lossless 15.3 foot long 0.2 inch diameter conductor over a perfect groundplane. Eznec gives the 1.821 MHz base impedance as .3004 -2169j. In other words, the antenna "looks like" .3004 ohms of load resistance in series with 40.32pF on 1821kHz. The return path for current is through the .3004 ohm resistance and 40.32pF capacitance, back to the ground of the antenna (it is a Marconi antenna).

Such a termination (load) would require a series inductance of 2169j (189.57µH) to cancel feedpoint capacitive reactance. A typical 190µH inductor

would be rather large, requiring somewhere around 53 turns when using a 4" by 4" form factor. One would expect a physically large inductor to have noticeable but very small displacement currents to the groundplane, when the small stray coil capacitance is compared to the 40.32pF termination capacitance. This raises two very important design guidelines:

- ***When installing a loading coil of substantial inductance in an electrically short antenna, sheetmetal and dielectrics should be kept away from the coil and areas of antenna above the loading coil. This would include dielectrics on or near the inductor, since the presence of dielectrics would increase undesirable capacitance.***
- ***When inductive reactance requirements are large, as when short thin "stingers" without hats are used above a coil, the coil form factor should lean more towards long and thin. Capacitances near the open end of the coil (high voltage end) should be minimized. This would be true even when the coil length increase results in a small reduction in mutual turns coupling, since the stray capacitance may result in a larger loss penalty than the slight increase in accumulated resistance from additional wire length.***

Efficiency

Efficiency in any antenna near earth is almost always dominated by ground related losses, short-height Marconi antennas are no exception. The overall effect of loading inductor Q and matching system losses are "diluted" or "swamped-out" by ground losses. Ground losses cause most systems to have greatly reduced sensitivity to inductor design.

The only consistently predictable factor in efficiency in fractional wavelength Marconi antennas with limited size ground systems is radiation resistance. Efficiency increases almost directly in proportion to radiation resistance.

Radiation Resistance and Power Radiated

Radiation resistance is probably the most poorly defined term used with antennas. The lack of clear definition creates errors and misjudgments when predicting antenna performance. If you wish more detailed information, this page contains information on radiation resistance. For the purposes of this discussion and to avoid pitfalls associated with using feedpoint impedance as radiation resistance, I'll use the same definitions Jasik, Balmain, and others have used. This definition is based on the IRE definition of radiation resistance being equal to the net or effective current causing radiation squared divided by the power radiated as EM energy, or Rr=Pr/I^2.

Using this definition, a folded dipole has a radiation resistance identical to a conventional dipole of the same physical dimensions (~70 ohms).

Radiation is caused by charge acceleration, there is no magic. The only thing affecting radiation resistance in a short vertical antenna near ground is current distribution over the linear area occupied by the radiation portion of the antenna. The general rules are:

Radiation resistance of a Marconi vertical in the maximum possible radiation resistance case for a given height (this is the case where current is uniform throughout the structure) is equal to 1580*(H/L)^2 where H equals height and L equals wavelength and both are expressed in the same units. Using degrees, we see a 10-degree tall antenna has a maximum possible radiation resistance of 1580*(10/360)^2 or 1580*.000772 = 1.22 ohms. This would apply even if the antenna is a vertical, DDRR, Fractal, or folded unipole with considerable top loading.

If current is triangular, radiation resistance would decrease by a factor of four to 0.305 ohms.

Power radiated is given by I^2*Rr

With 100-watts applied to a 10-degree tall antenna, net current in a lossless antenna with uniform current distribution would be 9.05 amperes. With triangular distribution, such as appears in a small diameter short base loaded whip, current would be approximately 18.1 amperes. We are in serious problems if the inductor reduces current along its length, since the only possible way to radiate 100 watts would be to have somewhere around 9 amperes of effective current integrated over the 10-degree vertical area of space for the radiator!

Ground Losses

All current flowing (or displaced) vertically into the antenna must equal current flowing out of the ground or counterpoise system. Even though ground losses are distributed losses, we must normalize all losses to the feedpoint in order to compare systems. There are cases where this will not always occur, causing us to falsely assume we have lower losses than really exist.

In this tutorial and comparison, I have normalized ground losses to the same point where radiation resistance is considered.

System Losses

(Measured data below of actual antenna given below was from 1995 data taken at a different location near Atlanta with a slightly different loading coil and antenna. There is a slight disagreement with current data. I left this all in so you can see the departure from measurements and models using 8 year old data.)

Base Loaded (Triangular Antenna Current Distribution) with no ground loss

Assuming we have a base-loaded antenna, and the operating frequency has a wavelength of 550 feet (around the 160-meter band), a 15.3 foot vertical would fit the above 10-degree value. Interestingly enough when we compare Eznec to formulas available in older (1950 vintage) engineering textbooks, we find radiation resistance predicted by Eznec is .3003 ohms while the triangular current estimate for the same height radiator is .305 ohms! This is an amazing degree of agreement, illustrating what we could do before modeling programs became available. (With perfect top loading, both Eznec and longhand calculations show approximately 1.2 ohms of radiation resistance.)

Assuming our 15.3 foot tall (10-degree) base-loaded antenna uses a coil Q of 200, the coil has 10.845 ohms of ESR. Total resistance with a perfect ground would be 10.85+.3= 11.15 ohms. Current into this system with 100 watts applied would be around 3 amperes, resulting in ~2.7 watts radiated and ~97.3 watts lost as heat in the inductor.

Doubling coil Q (400) would provide 5.73 ohms of base resistance with 4.18 amperes. Power radiated would be 5.2 watts, power lost as heat would be 94.8 watts. Efficiency does not quite double, changing from 2.7 to 5.2%. This results in a 2.8dB change in signal level.

Top Loaded (with no ground loss)

If we added a four-wire hat with 15-foot wires, current would no longer be triangular. While we wouldn't quite reach the optimum uniform distribution, current at the top would be about 78% of current at the antenna base. Feedpoint impedance would become 0.97 -551j, and the antenna would look like 0.97 ohms in series with 159pF.

Using a coil Q of 200, we would now have 2.76 ohms of inductor loss. Current becomes 5.18 amperes. Radiated power is 26 watts, while power lost as heat becomes 74 watts. Even in the perfect ground case, the change in efficiency caused by top loading is large. Top loading (with only the hat) results in 9.8 dB change in signal level when compared to the base loaded case when coil Q remains 200. Efficiency is 26%. The coil remains at ground level for easy matching and frequency change.

In this case current at each terminal of the loading coil would be essentially the same regardless of poor coil mounting techniques. In order to have significant current taper in the coil or in the bottom of the mast, shunt capacitance would have to be a significant compared to 160pF. The antenna's high input capacitance relaxes inductor and antenna mounting electrical requirements.

Base Loaded (high ground loss)

My F-250HD Super Cab pickup truck, when parked over open medium quality pasture land, has a ground resistance of about 20 ohms (normalized to the feedpoint) on 160 meters. Applying this ground loss to the base loaded antenna, the system has a feedpoint resistance of 20+.3=20.3 ohms. (This is reasonably close to actual feedpoint resistances measured with a similar operating antenna.) Adding coil losses, the system has 20.3+10.85=31.15 ohms. (**NOTE: Current coil is ~8 ohms ESR, 10.85 ohms is from ~8 year old data**) Current is sqrt (100/31.15) or 1.79 amperes.

This results in .96 watts radiated, and 99.04 watts lost as heat. Efficiency is now around .96%.

Substitution of a coil with a Q of 400 results in 25.7 ohms feed resistance, or 1.97 amperes antenna current at 100 watts. In this case efficiency is now 1.16% for 1.16 watts radiated. The change caused by doubling coil Q with high system ground losses is about 0.8dB, compared to almost 3dB in the perfect ground case! With a poor ground (in this case typical of a very large vehicle), a large change in coil Q produces little change in system efficiency.

Another Top Loaded (high ground loss) System Example (made prior to the EZNEC model above)

Using a large hat isn't practical in a moving mobile, although it could apply to fixed stations suffering with poor ground systems. When the hat is smaller, such as a mobile requires, the loading inductor can be moved higher in the system. Such a move would produce uniform current below the loading coil, with a current shape above the coil dictated by the construction of the upper portion of the antenna. My own mobile uses a six-foot diameter hat manufactured from stainless steel automobile antennas arranged in a spoke. I have no problems with wind or occasional obstructions. While unsightly, a modest hat is workable.

In order to keep the systems comparable I'll use the same radiation resistance provided by a large hat, but intentionally add high ground loss as a lumped resistance. This model ignores field losses near the antenna.

In this case we have 0.97 -551j as the inductor termination presented by the antenna. With ground losses normalized at 20 ohms and an inductor Q of

200, we have 20+2.76+.97 = 23.73 ohms of feedpoint resistance. Current is 2.05 amperes, and power radiated is 4.1 watts. Power lost is 95.9 watts.

Efficiency is 4.1%, a 6.3dB increase over a base-loaded triangular current system with the same lossy ground. This system is 8dB down from the same "top-loaded" distribution using a perfect ground.

When the system has significant fixed losses, increasing radiation resistance four times by top loading provides a similar dividend in system efficiency. At the same time a substantial increase in coil Q provides only minimal change in field strength.

Current Through Coil

Related pages:

Inductor spice model

There has been some speculation that current is high only in the first few turns of a loading inductor. Radiation comes solely from charge acceleration or current over spatial (in line) distance.

If any loading inductor shows substantial decrease in current over the length of the inductor, it is an absolute certainty that the inductor is poorly designed and that the system above the loading inductor is not contributing to system efficiency. The reason for this is very simple and straight forward. Any two-terminal component (even considering wire as a "component" applies) MUST have equal charges flowing into and out of each terminal. Voltages to other reference points can be different, but for every charge moving into one terminal a like number of charges MUST move out of the other terminal. Radiation, induction fields, and loss resistances have no influence on this rule.

In order to have any change in current, there must be an additional path or paths for charges. This path can be through leakage resistances, or through fictitious currents called displacement currents. Whatever the path, the total charge movements must be reconcilable. We simply can not have current "disappear".

The normal path upsetting "unbalancing" current into and out of each terminal in an inductor is provided by displacement currents through electric fields. As with any system, the amount of current flow is proportional to potential difference and impedance of the path. In order to shunt a substantial current out of an inductor, the potential difference between the ends of the path has to be high compared to the impedance of the path. The impedance of the stray path must also be reasonably low compared to the normal desired path.

Current diversion is problematic in very large inductors operated at (or very near) internal self resonance, when the self resonance is what we typically refer to as a "series-resonant" condition. This condition is common in plate chokes used in vacuum tube power amplifiers, where the system operates over many octaves of frequency range.

"Series resonances" inside components occur when distributed inductance forms a pair (or multiples of pairs) of "L" networks. The large series inductance from each end of a winding reacts with the small stray capacitance at the center, and forms a very high impedance transformation L network. The electrical potential at the center of the system becomes extremely high, and even the smallest amount of capacitance to surrounding objects will carry a substantial displacement current. The large displacement currents cause the terminal impedances to drop, and allow considerable current to concentrate in small areas of the component. At the same time, considerable voltage can be present. The normal result is arcing or destruction of the component, or failure of the system depending on the choke to operate.

Series resonance always occurs at a frequency higher than the self ***parallel*** resonant frequency of the component. A loading coil operating under such conditions would be required to have serious design errors to fall into this category, since the end termination capacitances should always be substantially higher than stray capacitance throughout the component. Failure to follow this rule would result in needless loss and reduced SWR bandwidth in an antenna.

The speculation or supposition that the first few turns of a loading coil carry most of the current is clearly untrue. In order to shunt current off, high series impedances would have to exist along with high stray shunting capacitance to areas removed from the radiator. Additionally, the remaining coil area connected to the top area of the antenna above the loading coil would have to present a high impedance to the area where current reduction occurs. This would never be the case, unless the top area of the antenna and loading coil are not resonant near the operating frequency.

A reasonable test for proper inductor and system design would be to remove the antenna above the loading coil, measuring system resonance. If resonance does not change substantially, the area above the coil is not correctly terminating the system. First-order self resonance of the inductor (parallel resonance), when removed from the system, should also be far above the operating frequency of the system. If self-resonance comes within three or four times the operating frequency range, the loading coil almost certainly will have needless performance shortfalls.

Conclusion

A normally functioning inductor has essentially equal currents throughout the inductor, loading coils are no exception. Any current difference requires a substantial current flow through undesired stray capacitances or leakage currents. Neither radiation or induction could change this, it is a basic rule of circuitry.

In a reasonably well-designed system, current into and out of the loading inductor should be substantially equal. Differences in current would indicate excessive and problematic undesired stray capacitance in the loading system design, or measurement errors.

Reduced sensitivity to coil Q is primarily a function of additional losses in the system, not reduction of current through the coil.

This page has had **visits since February 11, 2004.**

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[How to Check Signals] [Keyclicks] [MFJ-259B Calibration] [RF Noise Powerstroke Diesel]
[My Shack] [Noise] [Radiation and Fields] [Receivers] [Skin depth and Conductors]
[stacking broadside collinear] [Transmitter Splatter] [Verticals and Baluns] [W4AN]

E-ham

FT1000D Click Mod FT1000D NB mod

W4AN Trust Fund. Bill Fisher contributed a great deal to amateur radio. I hope we can give something back to help his family.

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Other Recommended Technical Info Sites:

http://www.eznec.com/index.shtml

http://fermi.la.asu.edu/w9cf/

http://www.seed-solutions.com/gregordy/Amateur%20Radio/W8WWV%20Experimentation.htm

http://www.ifwtech.co.uk/g3sek/

QRN USA Lightning Storm Map

For an instant check on USA mainland QRN click on this link or this link to a weather map of thunderstorms.

Some of this site's (W8JI.com) contents are listed below:

New Page Cebik Dipole and other short dipoles. Feedline and tuner analysis!!

Amplifiers

SB-221 Tank Voltages and the parasitic oscillation myth. The real explanations behind amplifier parasitic oscillations and alleged damage by HF amplifier parasitic oscillations. Look at voltage throughout the tank at frequencies between 1 and 300MHz, and decide for yourself. Related article at Arcs, what makes an amplifier arc?

Preamplifiers for receiving. Measurements of receiving preamps. Compare your favorite pre-amp (text corrections 4/27/03)

Arcs, what makes an HF amplifier arc? This explains causes of amplifier arcing in any amplifier, including the TL-922 amp, Ameritron amps, Heathkit SB220 amplifier, and any other HF

power amplifier for amateur or commercial service. Simple failures and design or operating shortfalls are often explained as parasitic oscillations

Bandswitch failures the real causes and cures of bandswitch arcing and failures in HF power amplifiers that are often attributed to poor parasitic suppression causing oscillation

How to load or tune your HF amplifier and why an amplifier arcs when tuned incorrectly

Inductors deals with antenna loading coils but also applies to amplifiers

Related page: Inductor spice model

Relays how to speed and sequence amplifier T/R relays and how antenna transfer relays affect amplifier failures

Tank Circuits two models showing tank voltages, currents, loss, and phase shift

Vacuum Tubes how they work in RF power amplifiers and why vacuum tubes arc in amplifiers Includes details on 3-500Z and 8877 3CX1500 A7 tubes. Learn reasons tubes fail in amps

Vacuum tubes, how they convert d-c to RF how RF amplifier tubes convert dc to RF power

VHF Parasitics and general stability. Is your amplifier unstable? If you own a TL922, SB220 series, 30L1, or other amps with grid "negative-feedback" you will want to read this! (rev 5PM Feb23/03) Spice models. The resistor think might be non-inductive might not be what you think. Includes internal photos of various resistors.

ANTENNAS and antenna related topics

New Page Cebik Dipole and other short dipoles. Feedline and tuner analysis!!

EH antenna: the E-H and CFA antennas. How they work. See how my early theoretical predictions and descriptions compare to actual field measurements of this most recent antenna theory

hoax! For measurements click here.

If you think ***baluns at the inputs of tuners*** are a good idea, go to this W7EL link and read the section on current baluns. You might be surprised!

Another suggested tuner link is ***W9CF.***

Balun and Core Selection: (mostly receiving) Contains information on core selection for transformers and baluns (it isn't a balum)

Balun, sleeve baluns: How a sleeve balun works and what is important

Balun, transmitting: Testing transmitting baluns and untrue folklore about choke, voltage, and current baluns. Dispelling myth that grabbing coax adequately tests a balun

Baluns: winding methods with toroid cores and debunking myth that a split winding improves performance

Baluns: Common-mode noise. How baluns can help reduce receiving system noise

BALUN TEST: New!!! Test data on transmitting baluns.

Beverages: Construction, Endfire Beverages, Broadside Beverages, Beverage Arrays for low band DX and other low noise receiving antenna arrays for 160 meters and other low frequency bands. Also see balun and core selection for Beverage transformers.

Combiner and Splitters: Contains information on Magic-T splitters and combiners, how they work and what they do

Detuning Towers: A tutorial with an important correction of the common error that we adjust for minimum current!! (revised slightly Feb 17, 2003)

Gain. Stacking (Broadside) and end-to-end (Collinear) gain. How it works, dispelling the 3dB myth.

Loading Inductors: A brief tutorial on loading coils (inductors) and inductor Q

Magnetic Loop Antennas: receiving

Mobile and Loaded Antennas: Loading coils, ground losses, and currents in the system. Related page: Inductor spice model

Power Stroke Diesel RF Noise: PowerStroke Diesel noise and how I corrected the problem in my 2003 F-250

Noise and common mode noise. How it gets into the receiving system Power line and other noise sources

Omega and Gamma Matching: Contains technical information on Omega and Gamma matches, impedance limits, component selection, component failures

Phasing, Crossfire: Contains information on cross-fire phasing and why cross fire phasing is superior for broad bandwidth low-band receiving arrays

Phasing Systems: A quick discussion of phasing methods

Polarization and diversity: Think you can have the best of the two worlds of vertical and horizontal? Think again!

Radiation and Fields: Electric field, magnetic field, Fresnel zone What the terms we use actually describe

Radiation Resistance: A revised (as of Feb 14, 03) tutorial on radiation resistance and how it is used and misused

Receiving and receiving antenna pages: Relative ranking of antennas for Topband DX plus various articles on low-band DX receiving antennas...including low band receiving antennas such as Beverage antenna and Beverage antenna construction. Elongated Loops and other loop receiving antennas such as K9AY, EWE, and flag antennas, small Topband or HF low noise receiving Verticals , my receiving antennas (with a demo), end-fire or cross-fire Beverages, common-mode noise in low noise receiving systems, and more.

Skin depth: The best explanation I have found

Transmitting: Contains information on my various transmitting antennas, including my eight

direction four-square

Traps: Measurements and models of traps, including Coaxial and Tribander Antenna traps

Receivers

Diversity reception "diversity" receiving using stereo along with some sound file examples

How I test receivers

Receiver Tests of some transceivers and receivers. Newly added 756PRO FT1000 MK V Kenwood TS870 (modified)

Receiver IM improvement **mods for FT1000** FT1000D

DX Sound files: What some signals sound like here

1000MP MK V....mods for FT1000MKV receiver IM and transmitter clicks

Key Clicks

Includes sound files of signals with excessive bandwidth, a technical description of what causes clicks, and mods for the FT1000(D) and FT1000MP.

What Causes Clicks....The technical cause and what we can do about clicks.

Keyclick-MP.... An analysis and patch for the FT1000MP 's infamous key clicks

Keyclicks 1000D.... An easy and cheap patch for the FT1000 and FT1000D

Keyclicks 1000MP MK V....mods for FT1000MKV receiver IM and transmitter clicks

Also see Radios.

Noise

Technical article about noise and receiving/ receiving antennas

Power Stroke Diesel RF Noise: PowerStroke Diesel noise and how I corrected the problem in my 2003 F-250

MFJ-1025/1026 Technical Information

Common-Mode Noise issues with feedlines, and how to avoid problems

MFJ-259B Alignment instructions and some technical information MFJ259B.

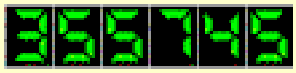
MFJ-1025 and 1026 phasing unit technical information.

Local ARES activities will be included later. We have a local repeater that covers a good bit of I-75 between Macon and Atlanta. Please give a call on 147.225(+600) if you are passing by.

Special thanks to my friend Bill Fisher, W4AN, who is now a silent key. It was Bill's encouragement and generosity that made this site available.

73,

Tom W8JI

This page has been viewed  *times since March 2003. Overall site hits are over 1.5 million for the past year!*

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NVIS - What It Is and How To Use It

By Patricia Gibbons - WA6UBE

NVIS , short for Near Vertical Incident Skywave, utilizes high-angle skywave paths between stations instead of ground-wave or surface-wave in order to communicate via HF radio. NVIS was originally evaluated by U.S. Army Forces in Thailand during the Vietnam conflict in the mid-1960's It was found that Mobile stations, using whip antennas bent parallel to the ground, could communicate more reliably with their base-stations. Signal strengths would be weaker using high-angle skywave but communications would be more reliable, less subject to fading, and consistent between stations. This was because the intervening terrain was less of an absorber of signals. Terrain obstructions between stations, such as hills, mountainous areas, jungle growth, built-up areas with tall buildings, no longer become path obstructions with stations when NVIS techniques are employed. For distances out to 400 miles between stations, one F-layer hop, at vertical angles of 45 degrees or higher are used. It is not necessary to have high power transmitters. Typical 100 watt power levels are fine. It is necessary that all stations on an NVIS radio network use antennas that are parallel to the ground and the frequencies used are chosen via a radio propagation prediction program in order to have best results.

Frequently asked questions about NVIS:

1: Isn't NVIS, when using a horizontal Dipole antennas, what amateur radio operators always have used? What is so different about it?

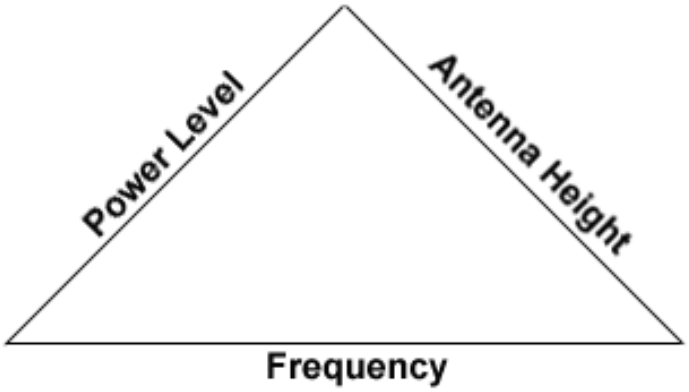
ANSWER: NVIS can be viewed more as a "Systems Concept" and not just what antenna to use. The concept of NVIS is to have reliable communications anywhere within an 800 mile diameter circle, in which your station would be located at the center of this circle.

2: What is the advantage of having the antennas close to the ground? I always thought that a radio antenna had to be as high in the air as possible.

ANSWER: Stations communicating via high-angle sky-wave may also be close enough to each other to receive a ground-wave or surface-wave signal. Stations receiving both a sky-wave and ground-wave will have heavily distorted received signals. This will include multi-path distortion because of the extreme difference in the lengths of the twp paths. Keeping antennas close to the ground will reduce the generation of a ground-wave signal.

3: Does NVIS work with low-power transmitters??

ANSWER: Yes. In fact, reliable communications between stations are based on three major factors. These can be viewed in triangle form as follows:



After much research and testing of antennas over the past 5 years, I have determined that the most important leg of the triangle is choice of correct frequency. Specific results will be discussed later on this page.

4: What are the typical frequency ranges used for NVIS?
ANSWER: Usually between 2.0 and 10 Mhz. Exact frequency is dependant on the degree of solar sunspot activity. The best choice of frequency is through the use of a propagation prediction program.

ACTUAL TEST RESULTS
Summer of 1990:

A field test of NVIS was planned between myself and Carl Sato, AA6CF. Carl was located in San Francisco, California. The plan was for my station to run RTTY on the amateur 40 meter band and for Carl to log the field strength of my transmitter at Carl's location. I would try different types of antenna arrangements. The plan was to have a "blind test" in that AA6CF would not know which antenna I would be on at any particular time. My station was parked in a local park in Morgan Hill, California, which is approximately 70 miles from Carl's station in San Francisco. My station was a mobile arrangement consisting of a restored military communications truck and used one of the following three antenna arrangements for the test:

1. 100 Ft horizontal long-wire at six feet above the ground and end-fed with an antenna tuner.
2. 15 foot military whip antenna tuned with an SGC model SG230 "smart-tuner" with the antenna in a vertical position. The base of the vertical antenna was 7 feet off the ground.
3. The same 15 ft military whip antenna but placed in a horizontal position behind the vehicle which allowed the horizontal whip antenna to be 7 feet above, and parallel to the ground.

The amateur 40 meter band was used. The mobile station was set up to transmit 50 watts of "mark-idle" signal. AA6CF would then tell me the signal strength of the signal received as I rapidly changed between one of the three antenna configurations as listed above.

TEST RESULTS:
The received signal strength from AA6CF on each of the three antennas were:

1. Horizontal wire antenna = S9
2. Horizontal Whip antenna = S9
3. Vertical Whip antenna = S8

Fall of 1990:

During the Boy Scout "Jamboree on the Air" (JOTA) event. My communications truck was set up at the San Jose Red Cross facility in their large parking lot.

Antenna Used: 1/2 Wavelength Dipole antenna, center-fed via 6 feet of ladder line and an MFJ-989C antenna tuner. Antenna was spaced 24 inches off the ground on orange traffic cones that were spaced six feet apart. The antenna was free from nearby obstructions as it was located in the center of a large empty parking lot.

Transmitter used: Yaesu FT70/G Paramilitary "Manpack" transceiver

Power used: 10 watts on CW

Frequency used: CW portion of 40 meter amateur band. 7025 - 7150 Khz.

Location of stations worked: All within a radius of 200 miles in the area of San Francisco, Emeryville, Sebastopol, Oroville, Woodland, Sacramento, and Dublin, California.

TEST RESULTS:
All stations reported my signal to be from "S8" to " 10 over S9" with exception of one station located in Plesanton California. The station in Pleasanton was using a Trap Vertical antenna. All other stations reported using horizontal or "Inverted Vee" Dipole antennas.

Spring of 1993:

Purpose of test: To compare two dipole antennas. One at 10-1/2 inches off the ground, and the other at 6 feet off the ground. Ground conditions were chosen to provide a "worst-case" as far as attenuation of the signal due to soil proximity. Soil conditions were extremely wet due to recent rainfall at a local park.

Antenna Used: Two types of antennas:

1. Dipole antenna at 10-1/2 inches off the ground supported by plastic tent stakes.
2. Dipole antenna at 6 feet off the ground and supported by short military telescoping masts.

Both antennas were balanced systems using ladder line from an MFJ989C antenna tuner.

Transmitter used: Yaesu FT70/G paramilitary "Manpack" transceiver

Power used: 10 watts on Single Sideband

Frequency used: Voice portion of 40 meter amateur band. 7225 - 7300 KHz

Location of stations worked: Wanted to find one station within NVIS range but far enough distance not to copy any ground-wave. Station worked was located in Menlo Park, California. This was a distance of 20 miles.

TEST RESULTS:
The station in Menlo Park, California reported my 10Watt PEP signal as:

1. "10 dB over S9" using the dipole antenna that was 6 feet off the ground.
2. "S8" using the dipole antenna that was 10-1/2 inches off the ground.

Post-Test equipment check: The Yaesu FT70/G was checked with a Motorola Model 2410 Communications Service monitor to verify the actual difference in meter indications for an "S" meter reading of "S8" and "10 dB over S9" The actual difference in level was shown to be "15 dB".

"Rules of Thumb" based on the above operations on the 40 meter band:

A. Assume a half-wave dipole at 1/4 wavelength above ground as a reference for comparison
B. A half-wave dipole at 6 to 7 feet off the ground will have an attenuation of approximately -4 dB
C. A half-wave dipole 10-1/2 inches off lossy ground will have a worst-case attenuation of approximately -20 dB
D. Assuming correct choice of frequency and a 10.7 cm solar flux value in the 200 range, a half-wave dipole at 1/4 wavelength above the ground would provide a 20 dB over S9 signal reading at the

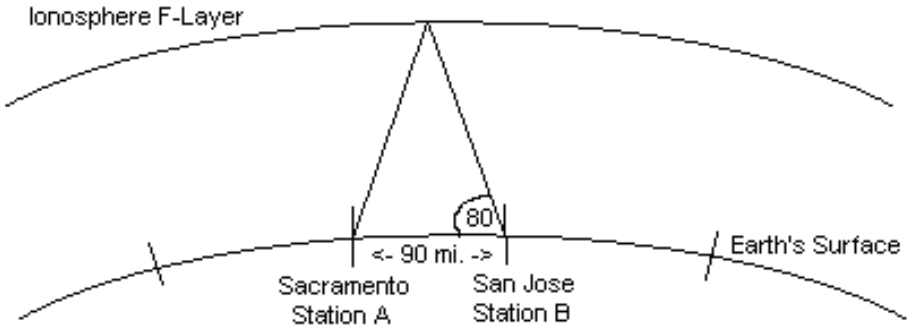
distant station when the transmitter has a power output of 100 Watts.

E. If the transmitting station uses antenna "B" above, the resultant signal strength would be: 16 dB over S9
F. If the transmitting station uses antenna "C" above, the resultant signal strength would be: S9

Based on actual documented tests between the station in Menlo Park and my station at the "San Jose Rose Garden" Municipal Park, This data tends to show that the antenna height above ground would not be the prevalent factor in establishing communications. For a 10 Watt radio to receive an "S8" signal report with a half-wave dipole at 10-1/2 inches off the ground on plastic tent stakes, it is apparent that the most important factor is proper choice of operating frequency. The Rose Garden tests were able to be done during a time when our Solar Sunspot activity produced a 10.7 cm solar flux value appropriate for the operating frequency used for this test. As of today's writing of this article (June 29th, 1995) report solar flux indexes are in the low to mid "70's".

Assume the crude drawing as follows:

Ionosphere "F-layer"



\ distance = 90 miles / Vertical radiation angle to reach from station "A" to "B" = 80 Degrees

FURTHER COMMENTS ON NVIS:

The U.S. Army did quite a bit of study toward the end of the Vietnam conflict on how to use HF radio more effectively and reliably. This effort was published in issues of "Army Communicator" magazine by Lt. Colonel David Fiedler starting in the early 1980's. Lt. Col Fiedler found that other countries, including German Ground Mechanized units of WWII, and the Soviet Union of today had implemented NVIS Since the summer of 1990, I have presented the NVIS concept at two west coast ARRL conventions and many local radio club meetings, as a way of publicizing this concept within the amateur radio community. Also, Ed Farmer, AA6ZM wrote a very extensive and well researched article on NVIS in the January 1995 issue of QST Magazine.

As a result of this work, Stanly Harter of the State of California Office of Emergency Services has taken a serious look at the value of HF communications for disaster communications. This is especially valuable for units like our California State Division of Forestry where operations in remote areas not served by the usual VHF and UHF mountain-top repeater sites could impact their ability to communicate effectively. Stanly Harter has also made recommendations for changes in HF antenna designs on their facilities used by State OES in order to effectively utilize the NVIS Concept.

In 1989, Just prior to the Loma Prietta earthquake here in the bay area, I had finished an equipment recommendation for the Director of our GSA-Communications Division with the City of San Jose. Included in the design of the radio equipment to by used by our San Jose Office of Emergency Services was an HF station which emphasized the use of NVIS high-angle skywave so that our center would have both County-wide and also solid, Northern and Central California coverage via HF communications. The Antenna consists of a 55 foot end-fed wire antenna mounted between two radio towers on the roof of our dispatch facility. At the top of one of these towers is an SGC "Smart-Tuner" which can then tune this horizontal wire on any frequency from 1.6 to 30 Mhz. The HF Radio used is a model RF3200 made by Harris/RF Communications Group. This HF station meets Part 90 rules for commercial type acceptance, and also covers any frequency from 1.6 to 30 Mhz. In an emergency, this station is capable of communications with State of California's Office of Emergency Services over their "Operation Secure" HF radio system.

Dr.Earl Stevens, Our Chief Radio Officer has his hand pointed to the ground to show relative scale.

Completed dipole antenna feed with ladder line installed 10 inches off the ground on plastic tent stakes.

10infeed shows the manner of connecting the antenna system to 3/4 inch ladder-line by the use of dual-banana plugs.

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NVIS: Near Vertical Incidence Skywave

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What is NVIS?

NVIS, or Near Vertical Incidence Skywave, refers to a radio propagation mode which involves the use of antennas with a very high radiation angle, approaching or reaching 90 degrees (straight up), along with selection of an appropriate frequency below the critical frequency, to establish reliable communications over a radius of 0-200 miles or so, give or take 100 miles. Although not all radio amateurs have heard the term NVIS, many have used that mode when making nearby contacts on 160 meters or 80 meters at night, or 80 meters or 40 meters during the day. They may have thought of these nearby contacts as necessarily involving the use of groundwave propagation, but many such contacts involve no groundwave signal at all, or, if the groundwave signal is involved, it may hinder, instead of help. Deliberate exploitation of NVIS is best achieved using antenna installations which achieve some balance between minimizing groundwave (low takeoff angle) radiation, and maximizing near vertical incidence skywave (very high takeoff angle) radiation.

As hams, we often faithfully follow the advice: get your antenna up as high as you can get it! We do this, and other things (like choosing antennas that have a low angle of radiation) in order to maximize the distance over which we can communicate. An antenna with a particularly high angle of radiation is often somewhat disparagingly referred to as a "cloudwarmer", the implication being that if the signal isn't radiated at a low enough angle, it's being wasted. For NVIS, we ignore all this traditional advice, and select instead techniques which will maximize not our DX, but our ability to reliably communicate with other stations within a radius of 0-300 miles.

Not just any old frequency will work for NVIS. Successful NVIS work depends on being able to select, or find (through trial and error), a frequency which will be reflected from the ionosphere even when the angle of radiation is nearly vertical. These frequencies usually are in the range of 2-10 MHz, though sometimes the limit is higher. The trick is to select a frequency which is below the current critical frequency (the highest frequency which the F layer will reflect at a maximum--90 degree--angle of incidence) but not so far below the critical frequency that the D and/or E layers mess things up too much.

Note: If you're already familiar with the ionosphere's role in HF radio communications, you may want to skip to the explanation of what's special about NVIS.

There are two main types of propagation at HF, known as "groundwave" and "skywave". Groundwave

propagation occurs when the receiving station is sufficiently close to the transmitting station, and is able to receive the portion of the transmitting station's signal which clings to the ground. The range of groundwave propagation varies with the type of antenna at the transmitting station, the characteristics of the ground between the transmitting station and the receiving station, and other factors. It can be anywhere from a few miles, to a few dozen miles. Distances beyond the range of the groundwave signal are covered by skywaves. Skywaves are the waves which radiate upward at some angle from the antenna, and (we hope) are reflected from the ionosphere, to return to earth further away.

The ionosphere is a high altitude region of the Earth's atmosphere which is composed of gaseous atoms which have broken into ions. The sun is the source of the ionizing energy, so the condition of the ionosphere varies with time of day, season of the year, the 11-year sunspot cycle, and the 27-day rotation of the sun. The layers of the atmosphere that effect radio propagation are the D, E, and F layers. I won't go into much detail in outlining their roles. If you're interested in this topic, entire books have been devoted to it. In a nutshell, it's the F layer which is usually involved in reflecting our signals back to earth, while the D layer absorbs our signals. The E-layer can either help, or hinder.

Long distance propagation of radio waves is usually achieved by their being reflected from the ionosphere, and returning to earth some distance away from their point or origin. (Follow along with the diagram if you wish.) Radio waves which have been radiated at a very low angle of radiation travel a long way before finally making it up to the ionosphere, strike the ionosphere at a very shallow angle (A) and return to earth far away from their point of origin (A'). As the angle of radiation goes up, the radio waves strike the atmosphere at a more moderate angle (B), and return to earth closer to their point of origin (B'). For any given frequency and current state of the ionosphere, there may be some maximum angle of incidence at which the ionosphere will reflect signals back to earth. Signals which strike the ionosphere at a higher angle of incidence than the current maximum will not be reflected at all, but will continue on out into space, instead (C). The area of the earth to which the reflection would have occurred will be in what we call the "skip zone" (unless it's close enough to the signal source to receive the groundwave signal). The skip zone is the region consisting of areas of the earth's surface which are outside the radius the transmitting station's groundwave will reach, and yet not far enough away to receive reflections of skywaves.

NVIS techniques concentrate on the areas which are often in the skip zone. The idea is to radiate a signal at a frequency which is below the critical frequency, at a nearly vertical angle, and have that signal reflected from the ionosphere at a very high angle of incidence, returning to the earth at a relatively nearby location. (See illustration.) Of course, no antenna radiates all its signal at exactly one angle, so the best we can get is a range of angles, ranging from perfectly vertical, to nearly vertical. The portion of the signal which is radiated at a vertical, or nearly vertical, angle reflects back to earth over some radius, which is determined by the lowest angle at which the antenna radiates much signal. Absorption by the D layer, and other factors, determine some minimum frequency below which the signal will no longer be usable, and usually some distance beyond which signals will no longer be usable.

For areas which are within the groundwave range of the transmitting station, the groundwave's presence

may interfere with the reflecting skywave. It may very well help, too. It all depends on whether the groundwave and the skywave arrive in phase, out of phase, or somewhere in between, and their relative strengths. If the groundwave arrives at about the same strength as the skywave, and the two are out of phase, the signal will disappear. Since the height of the ionosphere varies with time, phase alignment may drift from in phase, to out of phase, resulting in signal fading. For this reason, it's best to minimize groundwave radiation when using NVIS techniques, so that it will be less likely to interfere with the skywave.

Although this discussion has focussed mainly on the transmission of signals, there is a corresponding advantage of using NVIS techniques in reception, and a trick or two that are useful mainly for reception. The corresponding advantage is that if your antenna favors high angles for transmission, it will also favor high angles for reception. An antenna optimized for radiating at the high angles used for NVIS will also be optimized for receiving the skywaves which will be arriving at a high angle from the ionosphere. An antenna which does not radiate much groundwave signal will also probably not receive groundwave signals as strongly. When both stations are using antennas which are optimized for NVIS, the mode is favored both in transmission and reception, and those advantages add together, increasing the chances of reliable communication.

There is also an advantage inherent in the use of NVIS style antennas which applies only to receiving. The frequencies which are useful for NVIS (usually 2-10 MHz) are the same frequencies which are most susceptible to atmospheric noise. A major source of atmospheric noise is distant thunderstorms. Nearby thunderstorms are the worst, of course, but the noise from all possible sources adds together. Unless there is a nearby thunderstorm, most noise will be the sum of the noise from distant sources which are all propagated to the receiving antenna. Since an antenna optimized for NVIS is listening mostly to signals propagated from relatively nearby areas, and does not favor the reception of signals, static crashes, and other sources of noise and interference from more distant sources, it will not hear as much noise or interference as an antenna optimized for DX operation. The result is a better signal/noise ratio.

Often, taking measures which optimize a station's NVIS capabilities will drop the noise level substantially. Sometimes, the drop in noise can be maximized at the expense of some signal strength, and result in a communication circuit which has lower signal levels, but even more dramatically lower noise levels, for an even better signal/noise ratio than could be achieved by focusing only on maximizing signal levels.

So, selecting a frequency below the critical frequency, but not too far below it, and selecting an antenna which will radiate skywaves at a high angle, and minimize groundwaves and the reception of noise, are the essential tricks of establishing reliable communication in the 0-200 mile radius which is so often a challenge for HF operation.

What are the advantages and disadvantages of NVIS?

Among the many advantages of NVIS are:

- NVIS covers the area which is normally in the skip zone, that is, which is normally too far away to receive groundwave signals, but not yet far enough away to receive skywaves reflected from the ionosphere.
- NVIS requires no infrastructure such as repeaters or satellites. Two stations employing NVIS techniques can establish reliable communications without the support of any third party.
- Pure NVIS propagation is relatively free from fading.
- Antennas optimized for NVIS are usually low. Simple dipoles work very well. A good NVIS antenna can be erected easily, in a short amount of time, by a small team (or just one person).
- Low areas and valleys are no problem for NVIS propagation.
- The path to and from the ionosphere is short and direct, resulting in lower path losses due to factors such as absorption by the D layer.
- NVIS techniques can dramatically reduce noise and interference, resulting in an improved signal/noise ratio.
- With its improved signal/noise ratio and low path loss, NVIS works well with low power.

Disadvantages of NVIS operation include:

- For best results, both stations should be optimized for NVIS operation. If one station's antenna emphasizes groundwave propagation, while another's emphasizes NVIS propagation, the results may be poor. Some stations do have antennas which are good for NVIS (such as relatively low dipoles) but many do not.
- NVIS doesn't work on all HF frequencies. Care must be exercised to pick an appropriate frequency, and the frequencies which are best for NVIS are the frequencies where atmospheric noise is a problem, antenna lengths are long, and bandwidths are relatively small for digital transmissions.
- Due to differences between daytime and nighttime propagation, a minimum of two different frequencies must be used to ensure reliable around-the-clock communications.

What kind of antenna works well for NVIS?

Dipole

Once again, the dependable dipole antenna proves itself useful. One of the most effective antennas for NVIS is a dipole positioned from .1 to .25 wavelengths (or lower) above ground. When a dipole is brought very close two ground, some interesting things happen. The most interesting thing, from an NVIS perspective, is that the angle of radiation goes up. In the range of .1 to .25 wavelengths above ground, vertical and nearly vertical radiation reaches a maximum, at the expense of lower angle radiation (which we'd like to minimize, anyway, for NVIS). A dipole can be used at even lower heights, resulting in some loss of vertical gain, but often, a more substantial reduction in noise and interference from distant regions. Heights of 5 to 10 feet above ground are not unusual for NVIS setups, and some people use dipoles as low as two feet high with good results (relatively weak signals, but a very low noise floor).

Another interesting thing that happens with very low dipoles is that their feedpoint impedance goes

down. An acceptable SWR with 50 ohm coax is likely. Plan to bring your tuner along just in case, but you may get by just fine without it.

Yet another fortunate thing about low dipoles is that they are easily erected. Finding a tree which will serve as a support is often easy, and it's not hard to get a line in a branch which will suffice. Masts made of PVC tubing are practical at these heights. Very low dipoles can be supported by traffic cones with a notch cut in the top, or a simple tripod made from short sections of PVC pipe or wooden dowels, and bungee cords.

With the exception of the very lowest dipoles, most dipoles will gain an extra 2 db or so of vertical gain if you allow the center to droop a few feet. Allowing the center to droop means that the end supports don't have to be as sturdy, which makes installing a good NVIS dipole that much easier.

Inverted Vee

The dipole's close cousin, the inverted vee, is another good NVIS antenna, which can be even simpler to support. An inverted vee will work almost as well as a dipole suspended from a slightly lower height than the apex of the inverted vee, so long as the apex angle is kept gentle--about 120 degrees or greater. An inverted vee is often easier to erect than a dipole, since it requires only one support above ground level, in the center.

Counterpoises

The high angle radiation of a dipole (or inverted vee) can be enhanced by adding a counterpoise wire below it, about 5% longer than the main radiating element, to act as a reflector. The optimum height for such a counterpoise is about .15 wavelengths below the main radiating element, but when the antenna is too low to allow for that, a counterpoise laid on the ground below the antenna is still effective.

A knife switch at the center point of the counterpoise can be used to effectively eliminate the counterpoise from the antenna system. This technique is useful for using a dipole for NVIS and longer distances, too. A counterpoise is installed at ground level, or as high as the switch can easily be reached, and a dipole is mounted .15 wavelengths above the counterpoise. When the switch is closed, the vertical gain will increase, and the noise levels will drop. When the switch is open, lower angle gain will increase, improving the antenna's performance for non-NVIS use.

How do I select a frequency for NVIS operation?

The selection of a optimum frequency for NVIS operation depends upon many variables. Among the many variables are time of day, time of year, sunspot activity, type of antenna used, atmospheric noise, and atmospheric absorption. To select a frequency to try, one may use recent experience on the air, trial and error (with some sort of coordination scheme agreed upon in advance), propagation prediction software, near real-time propagation charts (available on the Internet) showing current critical frequency,

or even just a good educated guess. Whatever the strategy used for frequency selection, it would probably be best to be prepared with some sort of "Plan B" involving communicating through alternate channels, or following some pre-arranged scheme for trying all available frequency choices in a scheduled pattern of some sort.

In this discussion, some of my comments will assume that the reader's choice of frequencies is restricted to the amateur bands, but much of the discussion will be more general.

NVIS related links

http://www.ci.san-jose.ca.us/oes/races/hfradio.htm
http://www.gordon.army.mil/acd/tcs/hf/2418xtr2.htm
http://www.wr6wr.com/products/book_nvis.html

Instructions

Model HF6V

WARNING: DO NOT INSTALL THE ANTENNA WHERE ANY PART OF IT CAN COME INTO CONTACT WITH POWER LINES IN THE EVENT OF STRUCTURAL FAILURE, DURING INSTALLATION OR IN THE COURSE OF NORMAL FLEXING AFTER INSTALLATION FOR SUCH CONTACT CAN RESULT IN DAMAGE TO PROPERTY, BODILY INJURY OR EVEN DEATH!

WARNING: IN NO CASE SHOULD THE ANTENNA BE INSTALLED WHERE STRUCTURAL FAILURE OF ANY PART OF THE ANTENNA OR ITS SUPPORTING SYSTEM CAN ENDANGER PERSONS OR PROPERTY.

CAUTION! A GROUNDED ANTENNA WILL BE AT D.C. GROUND POTENTIAL! TO AVOID THE DANGER OF SHOCK CONNECT ALL STATION EQUIPMENT TO A GOOD EARTH GROUND. IT IS ALSO RECOMMENDED THAT ALL STATION EQUIPMENT BE DISCONNECTED FROM THE POWER MAINS BEFORE CONNECTING THE FEEDLINE TO THE ANTENNA. PLEASE CONSULT THE A.R.R.L HANDBOOK OR OTHER REFERENCE MANUALS FOR ADDITIONAL SAFETY PROCEDURES WHEN WORKING WITH ELECTRICAL EQUIPMENT.

NOTE: PLEASE READ ALL INSTRUCTIONS THOROUGHLY BEFORE PROCEEDING TO ASSEMBLY.

NOTE: HIGH PERFORMANCE BUTTERNUT VERTICAL ANTENNAS REQUIRE A RADIAL SYSTEM FOR ALL INSTALLATIONS.

Butternut offers three systems for installing vertical antennas:

- Model GRK Ground Radial Kit for ground mounting—160 thru 6 meter operation
- Model RMK-II Roof Mounting Kit for roof mounting—80 thru 6 meter operation
- Model CPK Capacitive Counterpoise Kit for compact installations below 25 ft (7.6 m) above the earth—80 thru 6 meter operation

Please refer to TECH NOTES—GROUND/RADIAL SYSTEMS, at the end of this instruction, for other mounting schemes and assistance in designing your own radial system.

REQUIRED TOOLS

Flat blade screwdriver and pliers. A 1/4", 11/32" and 3/8" nut driver will be helpful.

ASSEMBLY

Refer to the appropriate diagrams and proceed as follows:

1. Check to be sure that all parts are present.

2. Install tube w/insulator (A) per instructions packaged with mounting system or Tech Notes Ground/Radial Systems.

NOTE: A small packet of anti-seize/anti-oxide compound (Butter-It's-Not™) will be found inside tube w/insulator (A). This compound should be applied lightly to each tubing joint and to the inside of all clamps that must make good electrical contact with the tubing sections.

3. Locate tube (B) and tube (B1). Slide the insulator end of tube (B1) into the end of tube (B) with hole located 1/4 in (6.4 mm) from the end of the tube. Pass a # 8 x 1 1/2" screw through both parts and secure with a lock washer and hex nut.

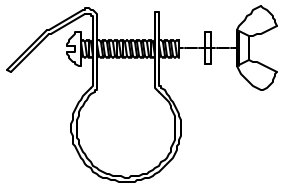
4. From the center of the insulator, measure downward to a point that is 13 in (33 cm) along tube (B) and make a pencil mark.

5. From the center of the insulator, measure upward to a point that is 9 3/8" (23.8 cm) along tube (B1) and make a pencil mark.

6. Locate coil assembly 80/40 meter (C) and slide the clamp at the outer end of the larger 80 meter coil over tube (B1), lowering the entire assembly until the middle clamp can be positioned around the insulator between tube (B) and tube (B1).

NOTE: The middle clamp may have to be pulled open slightly to pass the bolt that goes through tube (B1) and the insulator.

7. Position the center coil clamp of coil assembly 80/40 meter (C) in the center of the insulator between tube (B) and tube (B1). Pass a #10 x 1" screw through the clamp as shown. Secure with a flat washer, lock washer and wing nut.



NOTE: The outer tab of this clamp may be bent back slightly to provide clearance for the bolt, bending it back into place after assembly.

8. Stretch the 40 meter (smaller) coil on the coil assembly 80/40 meter (C) until the top of the upper clamp is even with the upper mark. Secure with a #10 flat washer, lock washer and wing nut.

9. Stretch the 80 meter (larger) coil on the coil assembly 80/40 meter (C) until the bottom of the lower clamp is even with the lower mark. Secure with a #10 flat washer, lock washer and wing nut.

10. Locate the capacitor assembly 80/40 meter (D) and install capacitor bracket 80 meter (D1) on the larger 200 pF capacitor using the installed screw.

NOTE: DO NOT USE EXCESSIVE FORCE OR OVER TIGHTEN THE SCREWS ON EITHER CAPACITOR AS YOU WILL DAMAGE THEM. DO NOT DROP THIS ASSEMBLY AS YOU MAY FRACTURE THE CAPACITOR'S CERAMIC SHELL.

11. Locate capacitor bracket 40 meter (D2) and install on the smaller 67 pF capacitor as above.

12. Install the above assembly onto the #10 screw protruding from the tab of the center clamp on the coil assembly 80/40 meter (C). Align capacitor bracket 80 meter (D1) alongside the larger 80 meter coil of coil assembly 80/40 meter (C). Secure with a #10 flat washer, lock washer and hex nut.

13. Attach the tab end of capacitor bracket 80 meter (D1) to tube (B) with capacitor bracket clamp and secure with # 8 x 1" screw, lock washer and a hex nut.

14. Attach the tab end of capacitor bracket 40 meter (D2) to tube (B1) as above.

15. Insert the un-slotted end of tube (E) into tube (B1) and secure with a # 8 x 1 1/2" screw, lock washer and hex nut.

16. Locate coil support tube 30 meter (O) and measure to a point 9 7/8 in (25.1 cm) down from the plastic insulator. Mark this point with a pencil.

17. Locate coil support tube 30 meter L bracket (O1) and place the tabbed end inside of the coil support tube 30 meter (O) securing it with a # 8 x 3/4" screw, lock washer and hex nut.

18. Place a #10 washer, lock washer and wing nut on the lower single clamp of coil/capacitor assembly 30 meter (P).

19. Place a #10 washer, lock washer and hex nut on both upper clamps of coil/capacitor assembly 30 meter (P).

20. Pass the lower single clamp of coil/capacitor assembly 30 meter (P) over the insulator end of coil support tube 30 meter (O) and slide the coil downward along the tube until the upper edge of the upper clamp is flush with the end of the insulator. Align the upper clamp with the coil support tube 30 meter L bracket (O1) and tighten the hex nut.

21. Stretch the coil until the bottom of the bottom clamp on the coil/capacitor assembly 30 meter (P) is even with the mark on coil support tube 30 meter (O) and tighten the wing nut.

22. Slide the remaining clamp from the above assembly over tube (E) and position it so the coil support tube L bracket (O1) is even with fourth turn, counting from the top of the 40 meter coil on the coil assembly 80/40 meter (C) and tighten the hex nut.

23. Hook the coil support tube 30 meter L bracket (O1) around the fourth turn of the 40 meter coil on coil assembly 80/40 meter (C). Secure with a # 8 x 3/4" screw, lock washer and hex nut.

24. Position wire clamp 0.875" 15 M w/insulator (K) around tube (F) and use a # 8 x 1" screw, lock washer and hex nut finger tight.

25. Insert the un-slotted end of tube (G) into the slotted end of tube (F) and secure with a # 8 x 1 1/4" screw, lock washer and hex nut.

26. Locate wire clamp 0.750" 15 M w/insulator (N) and position it around tube (G).

27. Insert the un-slotted end of tube (H) into the slotted end of tube (G) and secure with a # 8 x 1" screw, lock washer and hex nut.

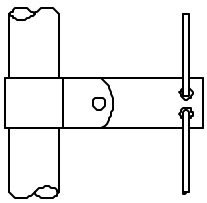
28. Locate wire clamp 0.625" 15 M w/insulator (M) and position it around tube (H).

29. Insert the un-slotted end of tube (I) into the slotted end of tube (H) and secure with a # 8 x 1" screw, lock washer and hex nut.

30. Position wire clamp 0.500" 15 M w/wire around tube (I) so the top edge is 13.5 in (34.3 cm) from the upper end of the tube.

31. Measure from the rivet of wire clamp 0.500" 15 M w/wire (L) to a point 11 ft 3 in (3.4 m) along the stranded wire and mark this point.

32. Pass the free end of the stranded wire from wire clamp 0.500" 15 M w/wire (L) through the small holes in wire clamp 0.625" 15 M w/insulator (M) and wire clamp 0.750" 15 M w/insulator (N) as shown.



33. Loop the end of the wire through the hole in wire clamp 0.875" 15 M w/insulator (K) sliding it on tube (F) until the mark on the wire appears. Wind the wire back on itself. Do not cut off the excess wire.

34. Line up wire clamp 0.875" 15 M w/insulator (K), wire clamp 0.750" 15 M w/insulator (N) and wire clamp 0.625" 15 M w/insulator (M) with wire clamp 0.500" 15 M w/wire (L) and tighten all clamps making sure the wire is moderately taut but not enough to cause the upper tubing section to bow.

ASSEMBLY

35. Place the protective cap on one end of tube (J).

36. Slide the uncapped end of tube (J) into the slotted end of tube (I) until only 25 in (63.5 cm) extends and secure with compression clamp small adjustable.

NOTE: In the following steps the antenna will be assembled and raised to its full vertical height. If the antenna is to be installed in an elevated position where it is unsafe or inconvenient to make in-place adjustments, the antenna may have to be installed in one piece. It will probably be necessary to raise and lower it and its supporting structure a number of times to arrive at the *ideal* adjustment on all bands. If so, every precaution should be observed in order to avoid possible contact with power lines and to prevent structural failure that can cause injury to persons or property.

37. Place the lower end of tube (B) through tube (E) over the insulator on tube (A) w/insulator. Line up the holes and secure it with a # 8 x 2" screw, lock washer and hex nut.

WARNING: AVOID POWER LINES!

38. Raise the assemble of tube (F) through tube (J) and slide the lower end into tube (E) fastening it securely with a # 8 x 1 1/4" screw, lock washer and hex nut.

39. Install coax 75 ohm matching (R) as shown placing the lug from the center conductor over the screw on tube (B) and the braid over the screw on tube w/insulator (A).

40. Place # 8 washers over each screw and install coil (Q) base matching. Secure with washers, lock washers and hex nuts.

NOTE: Attach radials and ground to tube w/insulator (A) using the remaining # 8 hardware.

WARNING: MAKE SURE THAT THE STATION EQUIPMENT IS CONNECTED TO A GOOD EARTH GROUND! DO NOT HANDLE CABLE CONNECTED TO STATION EQUIPMENT WITHOUT FIRST DISCONNECTING THE EQUIPMENT FROM THE POWER MAINS. YOU COULD BE ELECTROCUTED!

41. Connect coax 75 ohm matching (R) to any length of 50-53 ohm coaxial cable. Connector PL258 (S) is provided. Seal the connection with the small roll of Konnector-Kote.

CHECKOUT AND ADJUSTMENT

The dimensions and coil settings given above should produce reasonably low VSWR readings over the entire 10, 15, 20 and 30 meter bands and over at least 250 kHz of the 40 Meter band. Bandwidth on 80/75 meters should be at least 30 kHz for VSWR of 2:1

or less at the low end of the band and may be as much as 100 kHz at the high end of the band, depending on the efficiency of the ground system used, greater bandwidth being associated with lossy ground systems. It should be remembered that on those bands where the physical height of a vertical antenna is less than a quarter wavelength, the earth (or the resonant radial system in above-ground installations) will have a good deal to do with VSWR and antenna tuning, bandwidth and overall performance.

Low VSWR by itself does not mean that a vertical antenna is operating efficiently, and if low VSWR is obtained with no more than the usual *quick and dirty* ground connection, it most likely means the opposite. In general, poor operation or improper tuning of vertical antennas can usually be attributed to inadequate (or even reactive) ground systems or to other vertical conductors in the vicinity of the antenna. For these reasons it is suggested that the antenna be placed as much in the clear as possible and used with the best ground system that conditions permit. For a more complete discussion of the interrelationships between vertical antenna efficiency, bandwidth, VSWR, etc., a standard text such as the A.R.R.L. Antenna Book is recommended. See also the material included at the end of these instructions.

For adjustment purposes a simple VSWR indicator may be used. More accurate measurements may be made at the antenna (i.e., at the junction of the coax 75 ohm matching (R) and the main transmission line) than at the input end of the line, but the tuning conditions that exist at the transmitter will usually be of greater interest in that one's principal concern will be to couple power from the transmitter into the transmission line.

1. Determine the frequency at which VSWR is lowest on 80/75 meters. The coil setting given earlier should produce resonance and lowest VSWR at approximately 3700 kHz. To raise the frequency of resonance of the lowest VSWR, simply loosen the wing nut on the lower coil clamp of the coil assembly 80/40 meter (C) coil on tube (B) and stretch the coil a bit more. To lower the frequency, compress the coil. A 1 in (2.5 cm) change in the setting of this coil will produce a frequency shift of approximately 125 kHz.

NOTE: Remember that the antenna tunes very sharply in this range and that high values of VSWR may be encountered only a few kHz either side of the lowest VSWR readings, so it would be well to take VSWR readings every 25 kHz or so to avoid *running past* the frequency of resonance and lowest VSWR.

NOTE: To minimize interference to other stations and to avoid erroneous reading use only enough power to produce full-scale deflection of the meter in the *forward* or *r.f. out* position.

2. Once the proper coil setting has been found for the desired band segment, coil (Q) base matching at the base of the antenna may be adjusted for even lower VSWR. If earth losses are moderate to high a good match may be possible if coil (Q) base matching is left fully compressed; if earth losses are low (as with an extensive radial system) coil (Q) base matching may have to be stretched to twice its compressed length or more for a good match. In any case, a single setting for coil (Q) base

matching should suffice for operation over most of 80/75 meters provided the 80 meter coil is readjusted for each different band segment.

3. Determine the frequency of minimum VSWR on 40 meters. The coil setting given earlier should produce resonance and lowest VSWR at approximately 7150 kHz. The 40 meter VSWR and resonance curve may be shifted in the same manner as on 80/75 meters by changing the setting of the upper coil clamp of coil assembly 80/40 meter. On this band the setting is much less critical, and a 1 in (2.5 cm) change in the clamp setting will shift the VSWR curve approximately 80 kHz. Be sure to loosen the clamp around tube (E) that supports the 30 meter assembly and to reposition it as needed to avoid distorting the 40 meter coil.

4. Check VSWR on 20 meters. Tuning is quite broad on this band because the antenna is physically much taller than a quarter wavelength. To raise the frequency of the lowest VSWR, reposition the 30 meter assembly so that the coil support tube 30 meter L bracket (O1) can be replaced on the next lower turn of the 40 meter coil. Alternatively, to lower the frequency of lowest SWR, reconnect the coil support tube 30 meter L bracket (O1) to the next higher turn of the 40 meter coil. In some cases moving the tap point a full turn up or down may cause more of a frequency shift than is desired, in which case the entire 30 meter assembly may be rotated around tube (E) to permit adjustments of less than one full turn.

5. Check VSWR on 15 meters. The VSWR curve may be shifted upward or downward by changing the length of the stranded wire between wire clamp 0.500" 15 M w/wire (L) and wire clamp 0.875" 15 M w/insulator (K). To raise the frequency, simply shorten the wire by wrapping a longer *tail* back on itself and sliding the lower clamp upward to maintain tension. To lower frequency, feed more of the *tail* back through the hole in the insulator to increase the length of the wire between wire clamp 0.500" 15 M w/wire (L) and wire clamp 0.875" 15 M w/insulator (K). A change of 2 in (5.1 cm.) will shift the VSWR curve approximately 300 kHz.

6. Check VSWR on 10 meters. To raise the resonant frequency loosen the small hose clamp over the slotted end of tube (I) and slide tube (J) farther into tube (I); to lower the frequency, slide tube (J) farther out of tube (I) and retighten the hose clamp. A length change of 3 in (7.6 cm) should move the VSWR curve approximately 200 kHz.

7. Check VSWR on 30 meters. To raise frequency, loosen the wing nut on the bottom coil clamp of coil/capacitor assembly 30 meter (P), stretch the coil and retighten the wing nut. To lower frequency, compress the coil. A change of only 1/4 in (6.4 mm) will shift the VSWR curve approximately 100 kHz. Large changes in the setting of coil/capacitor assembly 30 meter (P) may affect 20 and 40 meter tuning, in which case it may be necessary to repeat steps 3 and 4. In general, the point at which the 30 meter coil taps on to the 40 meter coil will be the major factor in 20 meter tuning.

8. Adjustments for 40, 30, 20, 15 and 10 meters should have little or no effect on the previous adjustments for 80/75 meters, but a final VSWR check for this band should be made as in step 1 above.

NOTE: In above-ground installations it will usually be found that

resonance and lowest VSWR occur at slightly higher frequencies on all bands compared to ground-level installations. Therefore on 15 and 10 meters, where length adjustment is the means of getting antenna resonance, it is recommended that the length of the stranded-wire between wire clamp 0.500" 15 M w/wire (L) and wire clamp 0.875" 15 M w/insulator (K) be increased approximately 3 in (7.6 cm.) and that tube (J) be extended approximately 6 in (15.2 cm.) beyond the original dimensions given if any above-ground installation is contemplated. These are merely recommended preliminary settings, for it is impossible to indicate precise settings that will produce resonance or lowest VSWR at a given frequency in all installations.

In the preceding steps it has been assumed that the antenna has been installed in a more or less clear spot away from other vertical conductors such as TV antenna feedlines, towers and masts, and that a minimal ground system (or a system of resonant radials in the case of above-ground installations) has been installed.

If those fairly basic conditions have not been met it is likely that resonance and low VSWR will be impossible on some or even all bands. One should bear in mind that VSWR, even with a resonant antenna, will depend in large measure on local ground conductivity, height above ground in the case of an elevated antenna, the extent of the radial, counterpoise or other ground system used, and on other factors over which the operator may have little or no control. Fortunately, the evils of VSWR greater than unity have been grossly exaggerated in recent decades, and the only practical difference between a VSWR of unity and one of, say, 3:1 in the average case lies in the reluctance of modern equipment to deliver full power into lines operating at the higher VSWR without the help of a transmatch or other outboard matching device. Transmitters having so-called broadband solid-state output circuits (no tuning or loading controls) may be especially troublesome in this regard, whereas the older vacuum tube pi-network transmitters can usually be adjusted for maximum output over a tuning range where the VSWR does not exceed 2:l.

THEORY OF OPERATION

The first L/C circuit generates enough reactance to bring the whole HF6V to resonance on 80 meters allowing it to act as a 1/4 8 radiator. It also generates enough *capacitive* reactance to produce another discrete resonance at about 11 MHz. The second, 40 meter L/C circuit generates enough reactance to resonate the whole HF6V allowing it to act as a 1/4 8 radiator. In order to minimize conductor and I^2R losses an 80 and 40 meters where the antenna is physically shorter than a 1/4 8 and thus operates with lower values of radiation resistance, large-diameter self-supporting inductors and low-loss ceramic capacitors are employed. Where the height of the HF6V is slightly greater than a 1/4 8 on 30 meters, an L/C series tuned circuit taps onto the 40 meter coil for the extra inductance to pull the earlier 11 MHz secondary resonance down to 10 MHz. At the same time, a portion of the 40 meter coil is shorted out which allows the circuit to resonate on 30 meters The addition of this circuit also produces additional resonances at 14 MHz and 28 MHz. On 20 meters the entire radiator operates as a 3/8 8 vertical with much higher radiation resistance and VSWR bandwidth than conventional or *trapped* antennas having a physical height of 1/4 8 or less. Because the 20 meter radiation resistance will be several

THEORY OF OPERATION

times as greater as that of conventional vertical antennas, an electrical 1/4 8 section of 75-ohm coax is used as a *geometric mean* transformer to match the 100-odd S of feedpoint impedance on that band to a 50 S main transmission line of any convenient length. The HF6V operates as a slightly extended 1/4 8 radiator on 15 meters, a 1/4 8 stub decoupler providing practically lossless isolation of the upper half of the antenna on that band. On 10 meters the HF6V becomes a 3/4 8 radiator with considerably greater radiation resistance and efficiency than 1/4 8 trapped types.

ELECTRICAL AND MECHANICAL SPECIFICATIONS

Height (adjustable): 26 ft (7.9 m)
Shipping Weight: 14 lbs (6.3 kg)
Feedpoint Impedance: Nominal 50 ohms through included matching line.
VSWR at resonance: 1.5:1 or less all bands
Power rating: 2 kW PEP 75/80, 40, 20, 15, l0 meters; 500 W PEP 30 meters
Wind loading area: 2 ft^2 (.19 m^2)
Bandwidth for VSWR of 2:1 or less: entire band 10, 15, 20, 30 meters, 250-300 kHz
40 meters, 40-100 kHz 75/80 meters

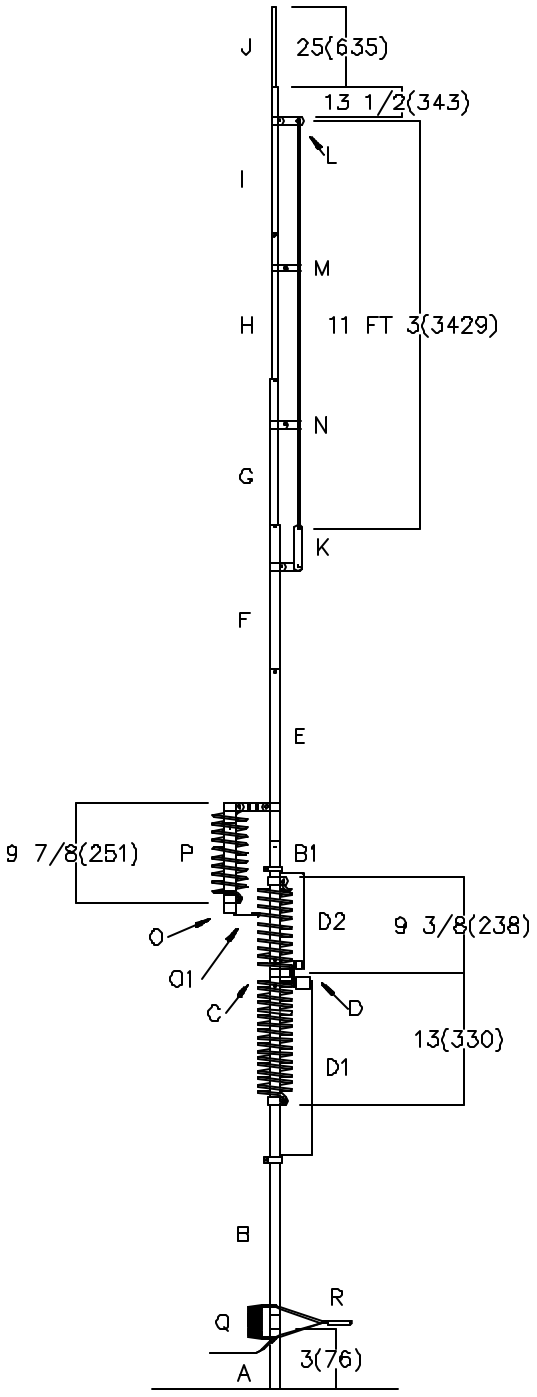
GUYING

The HF6V is designed to survive winds of up to 80 mph (129 kph) without guying in the absence of ice loading or heavy precipitation, but over a period of time it is to be expected that frequent or even constant flexing or vibration will reduce the chances for survival in winds that would not damage a newly installed antenna. Therefore in areas of frequent or heavy winds a set of short non-conductive guys should be used to reduce the stresses that wind loading will impart to the lower sections of the antenna. In this connection, it should be noted that light nylon twine is totally unsuitable as guying material because it has too much stretch per unit length, although the heavier sizes of nylon rope (or even sash cord) may be suitable if used in short runs. Polyethylene rope may be used, but because some grades tend to deteriorate fairly rapidly, periodic inspection should be made. A single set of guys placed just above the 30 meter circuit will contribute greatly to the stability and the longevity of the antenna, provided that the guys retain a slight amount of slack and do not come off at too steep an angle. At Angles of less than 45/ the guys begin to exert a downward compressive force on the structure that can be more of a threat to survival than lateral wind loading on an unguyed structure. Under no circumstances should guys be placed higher than one-third of the way up the antenna. The upper two-thirds of the HF6V has little more than its own weight to support, so these sections may be allowed to bend with the wind with no serious risk of damage. It is the lower third of the antenna that must support both the weight of the upper sections and the wind loading on them and are thus more likely to receive damage in severe winds.

PARTS LIST/PICTORIAL

NOTE: All dimensions are in inches(millimeters) unless otherwise noted

Code	Part No	Description	Qty
A	00278SZV	TUBE A W/INSULATOR 1-1/8(29) X 24(606)	1
B	00115BAV	TUBE 1-1/8(29) X 48(1216)	1
B1	00365SZV	TUBE B1 W/INSULATOR 1-1/8(29) x 12(302)	1
C	00145SZV	COIL ASSEMBLY 80/40 METER	1
D	290-07	CAPACITOR ASSEMBLY 80/40 METER	1
D1	00150BAV	CAPACITOR BRACKET 80 METER	1
D2	00220BAV	CAPACITOR BRACKET 40 METER	1
E	00123BAV	TUBE 1(25) X 48(1216)	1
F	00124BAV	TUBE 7/8(22) X 48(1216)	1
G	00125BAV	TUBE 3/4(19) X 48(1216)	1
H	00126BAV	TUBE 5/8(16) X 48(1216)	1
I	00127BAV	TUBE 1/2(13) X 48(1216)	1
J	00175BAV	TUBE 3/8(10) X 36(911)	1
K	00286RZV	WIRE CLAMP 7/8(22) 15M W/INSULATOR	1
L	00280RZV	WIRE CLAMP 1/2(13) 15M W/WIRE	1
M	00281RZV	WIRE CLAMP 5/8(16) 15M W/INSULATOR	1
N	00282RZV	WIRE CLAMP 3/4(19) 15M W/INSULATOR	1
O	00204SZV	COIL SUPPORT TUBE 30 METER 1-1/8(29) X 9(225)	1
O1	00176BAV	COIL SUPPORT TUBE 30 METER L BRACKET	1
P	00249SZV	COIL/CAPACITOR ASSEMBLY 30 METER	1
Q	00137SZV	COIL Q BASE MATCHING	1
R	290-08	COAX 75 OHM MATCHING	1
	00077JZV	# 8-32 X 3/4(19) SCREW	3
	00078JZV	# 8-32 X 1(25) SCREW	5
	00079JZV	# 8-32 X 1-1/4(32) SCREW	3
	00080JZV	# 8 LOCK WASHER	18
	00081JZV	# 8-32 HEX NUT	18
	00083JZV	# 8 FLAT WASHER	7
	00109JZV	# 8-32 X 2(51) SCREW	2
	00114JZV	# 8-32 X 1-1/2(38) SCREW	3
	00131JZV	#10-24 X 1(25) SCREW	2
	00132JZV	#10 FLAT WASHER	8
	00133JZV	#10 LOCK WASHER	8
	00134JZV	#10-24 HEX NUT	4
	00135JZV	#10-24 WING NUT	5
	00050DZV	KONNEKTOR-KOTE 1(25) X 8(203)	.05
	00089FZV	PROTECTIVE CAP 0.375	1
	00143BAV	CAPACITOR BRACKET CLAMP	2
	00144JZV	COMPRESSION CLAMP SMALL ADJUSTABLE	1
	00061SZV	BUTTER-IT'S-NOT	1
	00366IZV	INSTRUCTIONS HF6V	1



TECH NOTES—GROUND RADIAL SYSTEMS

MOUNTING TUBE INSTALLATION

When tube w/insulator (A) is ground mounted, it should be protected against corrosion if placed in concrete, damp acidic or alkaline soil. Asphalt roofing compound, polyurethane varnish or other sealant that protects against moisture may be used.

Concrete may be used in areas of high winds for greater strength, in which case the post may be twisted slightly during setting for easy removal later.

Tube w/insulator (A) must be installed in a hole approximately 21 in (53.3 cm) deep so that the upper end of the fiberglass insulator is approximately 7 in (17.8 cm) above ground level. Pack earth tightly around tube w/insulator (A) so that it remains vertical.

NOTE: HAMMERING TUBE W/INSULATOR (A) INTO THE EARTH MAY CAUSE THE INSULATOR TO SPLINTER. If the post must be hammered into the earth, protect the end of the insulator with a block of wood

NOTE: DO NOT USE U-BOLTS TO ATTACH TUBE W/INSULATOR (A) TO A MAST, TOWER ETC. U-BOLTS WILL EVENTUALLY CUT INTO THE TUBING AND WEAKEN THE INSTALLATION. If U-bolts must be used, place a larger diameter metal, such as the MPS Mounting Post Sleeve over tube w/insulator (A). Similar precautions should be observed when using TV style towers with locking bolts.

The RMK-II Roof Mounting Kit includes the MPS as well as the STR-II Stub Tuned Radial Kit.

GROUND MOUNTING

A vertical antenna in its simplest form, is electrically equivalent to one-half of a dipole antenna stood on end. When the antenna is mounted close to the ground, the earth below takes the place of the "missing" half of the dipole. If ground conductivity is fair to good, a short metal stake or rod may provide a sufficiently good ground connection for resonant and low SWR operation on the bands for which the antenna is designed. This basic arrangement is shown in figure 1.

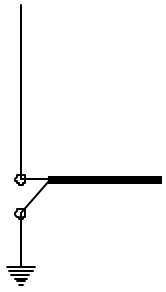


Figure 1

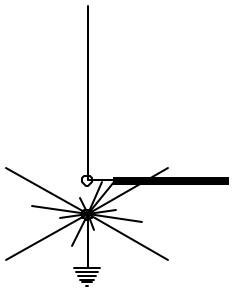


Figure 2

The way it works is that the capacitance between the vertical radiator and the ground causes *return* currents to flow along the earths surface back to the transmitter. If they have to come back along untreated lossy earth thy get back to the source greatly attenuated. This *return* loss is like a resistor in series with the antenna radiation resistance and will therefore affect the feed point impedance. In almost every case the efficiency of a vertical antenna will be greater if radial wires are used to improve ground conductivity as in figure 2. It's important to note that there's no point in cutting radials to any particular length when ground mounting because the earth will detune them anyway. All you want to do is make the surface of the earth around the antenna more conductive than it is ordinarily.

TECH NOTES—GROUND RADIAL SYSTEMS

If you can't copper-plate the backyard, the best approach is to run out as many radials as possible, each as long as possible around the antenna in all directions. Radials may be left on top of the ground however they should be buried for the sake of pedestrians and lawnmowers.

How long should radials be? A good rule is no shorter than the antenna is tall because 50% of your losses will occur in the first 1/4 8 out from the antenna. If you have more than a dozen radials, they must be longer to get the most out of them which is why the FCC specifies 113 wires each .4 8 for AM broadcast stations—the equivalent of a zero-loss ground plane. Obviously, for most ham work this would be overkill.

In some cases wire mesh (i.e. chicken wire) may be used as a substitute for radial wires and/or a ground connection, the mesh or screen acting as one plate of a capacitor to provide coupling to the earth beneath the antenna.

It should be noted that a ground rod is useful only as a d.c. ground or as a tie point for radials. It does little or nothing to reduce ground losses at r.f. regardless of how far it goes into the ground.

Bare wire, insulated, any gauge, it doesn't matter. The current coming back along any one wire won't amount to that much.

EFFICIENCY

The importance of reducing losses in the ground system can be seen from an examination of a vertical antenna's feedpoint impedance which at resonance consists of three components: antenna radiation resistance; conductor loss resistance; and earth loss resistance. An unloaded quarter-wave vertical antenna has a radiation resistance of about 35 ohms with negligible ohmic or conductor loss, but ground loss resistance may be very great if no measures are taken to reduce it, and in some cases ground loss R may even exceed the antenna radiation resistance. These three components may be added together to arrive at the feedpoint **impedance** of a **resonant** (no reactance) antenna. For the sake of illustration, assume that the ground loss beneath a quarter wavelength vertical antenna is 15 ohms, that conductor loss resistance is zero, and that the radiation resistance is the textbook figure of 35 ohms. The feedpoint impedance would then be 15 + 0 + 35 = 50 ohms, and the antenna would be perfectly matched to a 50 ohm coaxial line. Since the radiation resistance is an index of the amount of applied power that is consumed as useful radiation rather than simply lost as heat in the earth or in the conductor, the radiation resistance must be kept as high as possible in relation to the total feedpoint impedance for maximum efficiency. Efficiency, expressed as a percentage, may be found by dividing the radiation resistance by the total feedpoint impedance of a resonant antenna, so under the conditions assumed above our vertical antenna would show an efficiency of 35/50 = 70%. As a vertical antenna is made progressively shorter than one-quarter wavelength the radiation resistance drops rapidly and conductor losses from the required loading inductors increase. A one-eighth wave inductively loaded vertical would have a radiation resistance of something like 15 ohms and coil losses (or trap losses for multiband antennas) would be in the range of 5 ohms. Assuming the same value of ground loss resistance (15 ohms), the feedpoint impedance would become 15 + 5 + 15 = 35 ohms and

the efficiency would be 15/35 = 43%. From the above calculations it is clear that the shorter a vertical antenna must be the less efficient it also must be for a given ground loss resistance. Or to state the matter another way, more elaborate ground or radial systems must be used with shorter verticals for reasonable efficiency. If the ground loss of resistance of 15 ohms from the preceding example could be reduced to zero ohms, it is easy to show that the efficiency of our one-eighth wavelength loaded vertical would increase to 75%. Unfortunately, more than 100 radials each one-half wavelength long would be required for zero ground loss, so lower efficiencies with shorter radials must usually be accepted for the sake of convenience. In spite of their limitations, short vertical antennas over less than ideal ground systems are often more effective DX performers than horizontal dipoles which must be placed well above the earth (especially on the lower bands) to produce any significant radiation at the lower elevation angles. Verticals, on the other hand, are primarily low-angle radiators on all bands.

ABOVE GROUND (ELEVATED) INSTALLATIONS (rooftop, tower, mast. etc.)

The problem of ground loss resistance may be avoided to some extent by mounting a vertical antenna some distance above the earth over an artificial ground plane consisting of resonant (usually 1/4 8) radial wires. Four resonant radials are considered to provide a very low-loss ground plane system for vertical antennas at base heights of 1/2 8 or more. This arrangement contrasts favorably with the more than 100 radials for zero ohms loss resistance at ground level, and since 1/2 8 is only about thirty-five feet at 20 meters, very worthwhile improvement in vertical antenna performance can be realized, at least on the higher bands, with moderate pole or tower heights. At base heights below 1/2 8 more than four radials will be required to provide a ground plane of significantly greater conductivity than the lossy earth immediately below the antenna: even so, a slightly elevated vertical with relatively few radials may be more effective than a ground-level vertical operating over a larger number of radials if only because the former is apt to be more in the clear. Resonant radial lengths for any band may be calculated from the formula:

$$Length\ (ft) = \frac{240}{Frequency\ (MHz)}$$

Figure 3 shows the basic ground plane system for elevated verticals. Radials may slope downward as much as 45 degrees without any significant effect on operation or performance. Radials for different bands should be separated as much as possible and the far end of each radial insulated from supporting wires. Figure 4 shows a ground plane system that uses four resonant radials for 40 meters, another set of four for 20 meters, and a third set for 10 meters. A separate set for 15 meters is not ordinarily required because the 40 meter radials operate as resonant 3/4 8 radials on that band. At the lower heights the separate wires of this system may provide enough capacitance to ground to permit low SWR operation on 80/75 meters as well, but it is probable that at least one resonant radial will be required for low SWR on that band. It's important to note that cutting each conductor of rotator cable to a specific frequency will not work unless you separate it, angling each conductor away for most of its length because the longer ones will detune the shorter ones.

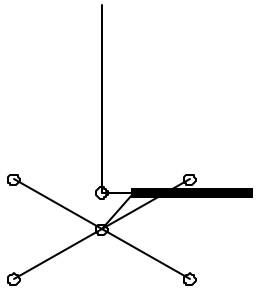


Figure 3

TECH NOTES—GROUND RADIAL SYSTEMS

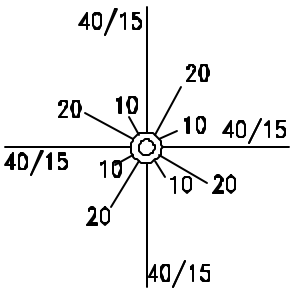


Figure 4

The 12-radial system of Figure 4 is a very good one, but it requires at least 12 tie-off points. Butternut has developed a multiband radial made of 300-ohm ribbon that resonates simultaneously on 40, 20, 15 and 10 meters. Four such radials offer essentially the same ground plane performance as the system of Figure 4 but require only 4 supports. These multiband radials plus additional wire for an 80 meter radial are available separately (our STR-II kit) or as part of the Butternut roof mounting kit (RMK-II).

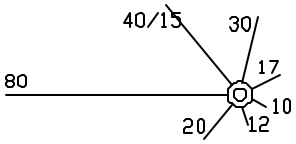


Figure 5

There are times when physical restrictions will dictate the use of fewer than four radials, and at least one manufacturer recommends 2 radials per band, the radials for each band running 180 degrees away from each other. A simpler (and no doubt less effective) system is shown in Figure 5. Since only one resonant radial is used per band the antenna will radiate both vertically and horizontally polarized energy, and the pattern will not be completely omnidirectional. For true ground plane action and predominantly vertical polarization no fewer than three equally-spaced radials should be used.

Figure 6 illustrates the construction of a multi-band radial which is resonant on 40, 20, 15 and 10 meters. Good quality 300 ohm TV ribbon lead should be used (velocity factor is critical), and the conductors should employ at least one strand of steel wire to support the weight of the radial. Four such radials will be the practical equivalent of the system shown in figure four for operation on 40 through 10 meters.

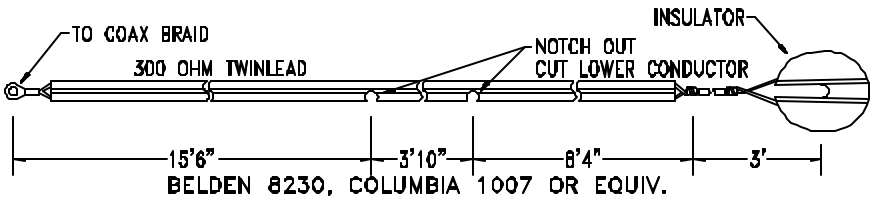


Figure 6

Regardless of the number of radials used in either elevated or ground level systems, all radials should be attached to the ground connection at the antenna feedpoint by the shortest possible leads. An elaborate radial system at ground level, for example, cannot be used with a vertical antenna on a rooftop or on a tall tower, for the length of the ground lead would effectively become part of the antenna, thus detuning the system on most or all bands.

METAL TOWERS AND MASTS

If a metal mast or tower is used to support a vertical antenna all radials should be connected to the mast or tower at the ground connection of the antenna feedline. This is because one of the functions of a resonant radial is to detune a supporting metal structure for antenna currents that might otherwise flow on the structure and thus turn the vertical antenna system into a vertical long wire with unwanted high-angle radiation.

OTHER MOUNTING SCHEMES

In cases where a resonant vertical antenna may neither be ground mounted nor used with an elevated ground plane, operation may still be possible if connection can be made to a large mass of metal that is directly connected or capacitively coupled to the ground, e.g., central air

conditioning systems or structural steel frames of apartment buildings. Some amateurs have reported good results with vertical antennas extended horizontally or semi-vertically from metal terraces which serve as the ground connection. Alternatively, a quarterwave vertical may be window mounted if a short ground lead to a cold water pipe or radiator can be used. If a long lead must be used, tuned radials may be required for resonance on one or more bands. Great care should be exercised in such installations to avoid power lines and to keep the antenna from falling onto persons or property.

MOBILE HOME AND RV INSTALLATION

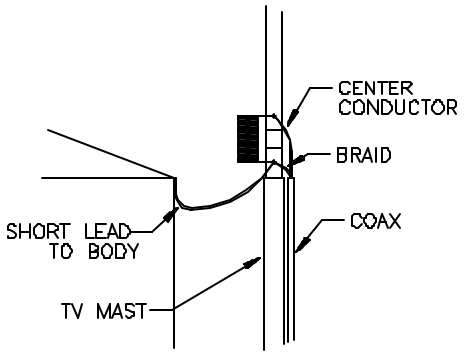
The principles of vertical antenna installations for use on mobile homes or RV's are the same as for other installations, and they all boil down to two main considerations. The first is that of erecting the vertical in the clearest possible spot, away from obstacles (including the MH or RV) that can interfere with radiation from the antenna. The second is that of installing the beat possible ground system beneath the antenna in order to minimize losses from r.f. currents flowing in the earth below the antenna. Fortunately, the metal bodies of both MHz and RVs can be used as highly conducting ground planes for vertical antennas in exactly the same way that automobile bodies, etc., provide the ground system for shorter vertical antennas for mobile operation. The metal body of an automobile, MH or RV may be viewed as one plate of a capacitor. Since the surface area of even a small automobile is quite large and in close proximity to the earth, its body is tightly coupled to the earth below and may be considered simply as an extension of the earth itself—a kind of hill as far as radio frequencies are concerned, but one having higher conductivity than the earth itself. RVs and especially MH having much greater surface area, will therefore provide a more extensive and effective ground system than a large number of radial wires occupying the same space as the MH or RV.

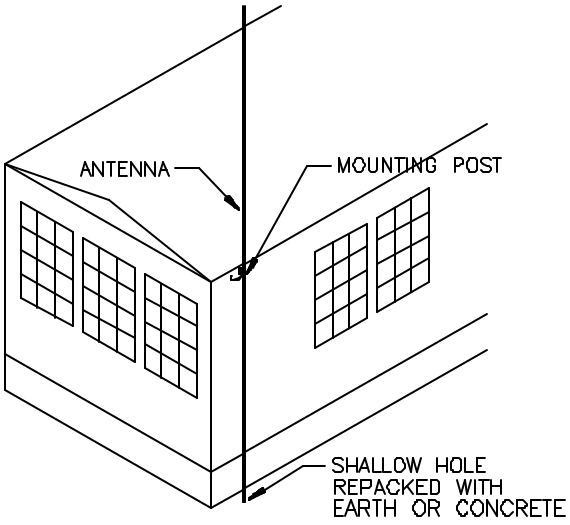
As in mobile installations, a vertical antenna may be mounted almost anywhere on the body of the vehicle or MH and made to operate with reasonably low VSWR, but it is generally considered that the best possible location for a mobile antenna is in the middle of the roof of the vehicle, i.e., at the center of the vehicle's ground plane and at a point where the antenna will not be in the "shadow" of any part of the vehicle. It is not usually convenient, or even practical to install a relatively tall vertical on the roof of an RV or MH for any number of reasons, so the next best procedure would be to install a vertical antenna with its base at the same level as the roof, preferably near the middle of one of the longer sides. The exact way in which this may be done is a matter of convenience, but a short mast extending from ground level to the roof of the MH and RV and placed alongside the building or RV would provide a stable and sturdy support with a minimum of mounting brackets and other modifications to the RV or MH. For portable operation such a mast could simply be lashed alongside the RV with the base in a shallow hole in the ground for additional support, and there would be no harm in extending the mast a few inches above the roof level to permit attachment of ropes which could be used to hold the mast firmly against the side of the vehicle and to prevent side sway.

This system has been used successfully with various types of RVs, travel trailers and even passenger automobiles during portable operation. For "L" shaped mobile homes a vertical antenna should be placed in the corner of the "L" so that the metal roof will provide groundplane coverage over 270 degrees.

TECH NOTES—GROUND RADIAL SYSTEMS

In all cases the base of the vertical antenna should not be more than a few inches away from the MH or RV so that the shortest possible lead may be run from the ground connection of the antenna to the metal body, as the length of this ground lead will effectively lengthen the antenna itself on all bands, and detuning can occur in some cases. A good electrical connection between the body of the RV or MH and the antenna is important, and in the case of mobile homes it would be a good idea to make sure that good electrical contact exists between the different parts of the metal body. Discontinuities can often lead to the production of harmonic radiation and TVI. The essential circuit connections are shown in the diagram above.





For permanent installations the bottom of the mast may be set deeper in the ground, and concrete may be used for greater strength and stability. The upper portion of the mast should be securely attached to the side of the building. Steel TV mast sections are readily available in lengths of ten feet and the mounting posts of Butternut HF verticals will slide into those which have an outside diameter of 1 1/4 inches and a wall thickness of .058 inches. Other vertical antennas may use different mounting techniques and requirements, so be sure to select a mast that will be suited to the particular situation. The main point to keep in mind is that the mast should not extend more than a few inches above the level of the roof so that the ground lead may be kept short.

LIGHTNING PROTECTION

Modern solid state amateur equipment is particularly vulnerable to damage from lightning or static induced transients that may appear on transmission lines, and conventional air-gap lightning protectors may provide no real protection at all for solid state gear. A line of very effective lightning and static protectors has been developed by ALPHA DELTA COMMUNICATIONS, P.O. Box 571, Centerville, Ohio 45459, for use with solid state equipment, and since these devices feature much faster transient discharge times than earlier designs, they should be investigated for possible use with all vertical and other antenna systems.

TROUBLESHOOTING

Check out your installation again, looking for loose connections and checking all dimensions. Then refer to the list of possible symptoms below:

Symptom: Few or no signals heard: bands seem *dead*, SWR is very high.
Look for: Open or shorted feedline, open or shorted matching line, broken connection at base of antenna (feedpoint).

Symptom: High SWR on 20 meter; other bands OK.
Look for: Missing matching line. Antenna not properly tuned. 20 meter radials not present or wrong length. Consult instructions for tuning and radial information; install matching line *RG-11 75 ohm coax, 11 ft 4 in (345.4 cm) if solid dielectric, 13 ft 6 in (411.5 cm) if foam type*.

Symptom: High SWR on some bands, but signals heard on all bands (conditions permitting).
Look for: Missing or defective radial system. Install as per instructions and check connections to radials and ground system. Keep this connection 6 in or less.

Symptom: High SWR on one band when antenna is roof-mounted. Radials are in place, but antenna will just not tune.
Look for: Radials of wrong length or running close to metal rain gutters or roof flashing. Tune radials and/or reroute them away from metal.

Symptom: Tuning is *sharp* with narrow bandwidth on 80 meter (and 160 meter if TBR-160-S is in place).
Look for: Normal condition. The total length of the antenna represents such a small percent of a wavelength on these bands that sharp tuning is a normal condition.

Symptom: Antenna was installed on the ground and tuned OK, but tuning changed over a period of weeks or months.
Look for: Antenna installed over poor ground system. Ground conditions have changed, causing shift in resonance. Install radial system as per instructions. Check connection to radial system. When you see this problem, you may assume that a ground rod without a radial system is not enough.

Symptom: Resonant point changes during wet weather.
Look for: Normal condition.

Symptom: Insulation arcs over between 80 meter and 40 meter coils damaging fiberglass.

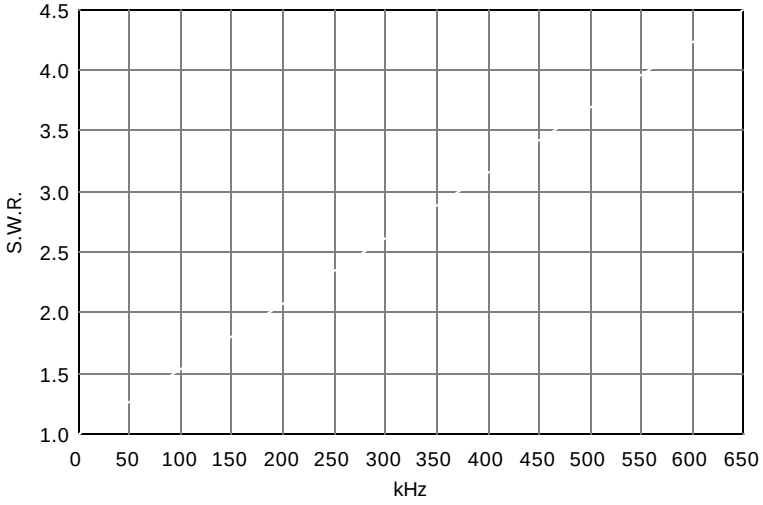
TROUBLESHOOTING

Look for: Operation at high power levels in areas where salt or pollution deposits have built up on the insulators. The cure is to keep insulators clean through routine maintenance.

Symptom: Intermittent operation. SWR jumps up and down suddenly, and reception is also intermittent.
Look for: Loose connections in the feedline or matching line (if used). Bad relay in rig. Bad antenna switch or connecting cable. Broken or corroded connections at the feedpoint. Bad radial/ground connection. Radial or antenna contacting metal when wind blows. Loose hardware on the antenna. Check and secure all connections.

Symptom: Antenna displays generally degraded performance after long period of time.
Look for: Lack of routine maintenance. Coax may be waterlogged or damaged. Build up of salt or pollution deposits on insulators and capacitors. Radial system corroded or rotted away. Owner must do routine maintenance at intervals, according to local conditions.

Symptom: SWR is OK on 75 meter, but goes up gradually when high power is applied. This is accompanied by heating of 200pF capacitor.
Look for: Bad ceramic capacitor. Replace.



Symptom: Antenna doesn't tune 80 meter or 160 meter, even though radials are in place and of proper length.
Look for: Antenna far out of tune; operator has not followed systematic tuning procedure. Start with suggested settings in instructions. Make an SWR chart to determine point of resonance. Adjust coils *carefully!* Remember, tuning, is *sharp* on these bands, so it is easy to pass the resonant point, then assume erroneously that the antenna isn't tuning.

BEFORE you call the manufacturer for help, please double check your installation, including all connections and dimensions. Tune carefully and systematically. Have SWR curves available. Be prepared to describe your installation in detail.

LIMITED WARRANTY

Butternut Manufacturing Co. warrants on the terms hereof, to a Customer who has purchased a Product from a Seller, for a period of one year from the date of the purchase, that the Product was not Defective, but this warranty is void if the Product has been subjected to improper or abnormal installation or usage, or a serial number on the Product has been defaced or removed.

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NO SELLER WILL BE LIABLE FOR ANY LOSS, INCONVENIENCE OR DAMAGE, INCLUDING DIRECT, SPECIAL, INCIDENTAL OR CONSEQUENTIAL DAMAGES RESULTING FROM THE USE OF OR INABILITY TO USE A PRODUCT, WHETHER THE LIABILITY WOULD RESULT FROM BREACH OF WARRANTY OR UNDER ANY OTHER LEGAL THEORY.

For instance, this warranty does not cover damage to or caused by an antenna (a) by reason of the antenna acting as a lighting rod, (b) by reason of corrosion or strain from exposure of an antenna to wind or weather, (c) from improper assembly, installation or use of an antenna, or (d) from failure periodically to inspect and maintain an antenna and its installation. The Customer is responsible to insure that installation and use of an antenna complies with applicable laws (such as zoning laws) and regulations (such as condominium regulations).

SOME LAWS DO NOT ALLOW THE EXCLUSION OF IMPLIED WARRANTIES, AND IF THESE LAWS APPLY, THEN ALL EXPRESS AND IMPLIED WARRANTIES ARE LIMITED IN DURATION TO SUCH ONE-YEAR PERIOD. NO WARRANTIES OF ANY KIND APPLY AFTER THAT PERIOD.

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Radiation&fields

[Home]

page moved to Radiation and Fields

Radiation Resistance

[Home] [Up]

Related pages on Antennas , radiation and fields , mobile and short verticals

My 2004 Dayton Hamvention Power Point presentation on Small Verticals can be downloaded here... DAYTON 2004.

The main points when dealing with small antennas are:

There is no magic bullet or magic cure to make a small antenna act like a large one. It all comes down to current distribution over linear distance.

Small antennas require extraordinary care to obtain high efficiency.

How do we make a *small* antenna as efficient as possible?

- First, we make current as uniform as possible over the length of the antenna.
- Second, we use low loss loading such as optimum form (size, length, and diameter) loading coils.
- Third, we make the antenna as large and straight as possible in a line. We don't fold, bend, zigzag, or curve the antenna especially in the high current areas.
- We keep the high voltage points (the open ends) away form other things (like lossy earth), and the high current areas away from other large lossy conductors.
- Most important, we keep current as high as possible throughout the length of the antenna by using as much capacitance as possible at the antenna ends.

What this does is maximize radiation resistance (while at the same time minimizing loss). The text below explains how radiation resistance and loss interact.

Radiation Resistance

Radiation resistance is probably the most abused and misused term in antennas. The reason it is so often misused is the lack of clear definition. When a term has several nebulous meanings and uses, it is only natural that misuses will appear. The lack of a firm well-accepted definition allows meanings to slip from one application into another, where a totally erroneous conclusion can be drawn from what otherwise would be a good formula!

Common Uses

There are two commonly used "correct" meanings of radiation resistance and one totally incorrect use. The "correct" uses are:

- The resistive part of an antenna's feedpoint impedance that is caused by radiation from the antenna
- The total EM power radiated in all directions divided by the square of net current causing the radiation

Neither of the above definitions include loss resistances of any type! The moment loss resistance is included, we have the third commonly used (but totally useless) definition. This definition, which includes losses, must be considered incorrect because it is comprised of resistances that have nothing to do with radiation. The misused definition is:

- The real (or resistive) part of an antenna's feedpoint impedance including loss related resistances.

The correct name for the third "radiation resistance is actually the antenna ***feedpoint resistance***, not radiation resistance!

Of the above ***good*** definitions, the first definition is the most commonly abused through mistake. The second definition is an IRE definition (albeit a good one that never caught on). In every case I've seen, it is the second good definition that always provides the most direct and useful answer.

Examples of Misuse

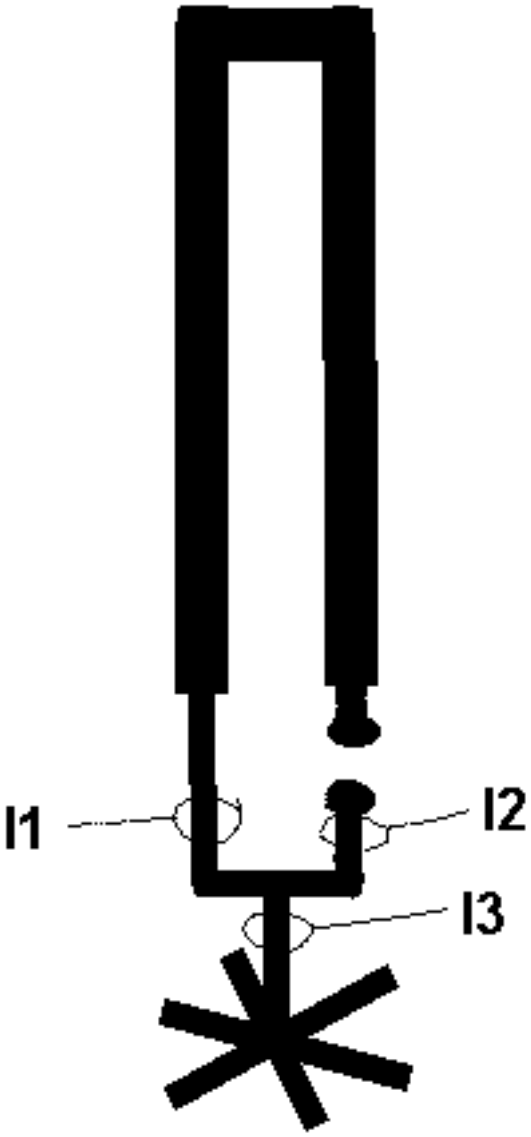
Folded Monopoles

Folded monopoles probably provide the best examples of misuse of the term radiation resistance. Quite often, in discussions of folded monopoles, claims are made that multiple drops increase the radiation resistance and lower losses. The justification for this incorrect claim is the folding raises

radiation resistance, and % eff = 100 * Rrad/(Rrad + Rloss) .

What folded monopole fanatics forget is that all losses must be normalized to the point where radiation resistance is taken, otherwise the efficiency formula won't mean a thing!

Let's look at what actually happens in a folded element, and use it to understand how the poor definition of radiation resistance causes the misunderstanding.



Consider the unipole above. Lets assume we short the open terminals, and feed it as a normal Marconi vertical with a feedpoint at the point where we measure I3. I3 is ALWAYS the vector sum or in-phase combination of currents I1 and I2. With continuity through each leg, I1 and I2 share all of the ground current. This happens regardless of where the feedpoint is located in the lower portions of the antenna.

With 1/4wl height and a reasonable element diameter, the radiation resistance (fed as a traditional monopole) would be about 36-ohms.

Assume ground loss, normalized to the point where we measure I3, could be represented by 14 ohms. Applying 500 watts would make current I3 equal 3.16 amperes. **Power loss in ground resistance would be I^2R, or 3.16^2 times 14, about 140 watts.** Feedpoint resistance would be 50 ohms. Feedpoint power, as a check, would be 3.16^2 times 50 ohms or 500 watts. With equal diameter legs, that current would divide and 1.58 amperes would flow in each upper leg.

Let's use the formula %eff = 100*Rrad/(Rrad+Rloss). We have 36/36+14 = .72 so the result is 72% efficiency, 28% loss. **28% loss times 500 watts is 140 watts in ground losses.** This matches the other method just above.

Opening the terminals and feeding as a folded unipole, half of the radiator current is in I1 while the other half is in I2. Current is halved to 1.58 amperes at the feedpoint and power remains the same. The feedpoint resistance now becomes 200 ohms. We can confirm this with I squared R, or 1.58^2 *200=500 watts. It all works out great so far!

Now let's MIS-use the same efficiency formula, like Orr does in his Handbook and others do in various places. We would have 200/200+14 = .9346 or 93.46 % efficiency.

We know we still have 3.16 amperes flowing as I3, and we know ground resistance is still 14 ohms (normalized to the point where I3 is measured). I-squared-R losses are 3.16^2 * 14 = 140 watts! We have exactly the same power loss.

Let's transform the ground loss value that was normalized at 14-ohms where I3 is measured to the feedpoint by the same impedance multiplication as the feed resistance, or 1:4. We'd now think ground resistance would be 4*14 = 56 ohms. 56 ohms of the feedpoint resistance is loss. Trying that same efficiency formula, we get:

144/144+56 = .72, or 72% efficiency!!! Now everything checks out fine.

The Common Mistake

Orr and others have used the first definition of radiation resistance, the portion of the terminal

resistance of the feedpoint responsible for radiation. Unfortunately ***they failed to normalize ground losses to the same point where the radiation resistance was taken!***

We can not use a formula that is based on everything being normalized to one point, unless we actually do that for every term in the formula!

There is no change in efficiency when the NET radiator current remains the same, and when ground current remains the same. It is pathological engineering to think otherwise.

Using The Second Definition

If we use the second IRE definition of radiation resistance, where the effective current causing radiation is compared to power radiated, we find nothing changes. A folded dipole or monopole has the same radiation resistance as a regular dipole or monopole the same size, and a small loop has the same radiation resistance regardless of turns.

The magic vanishes along with the incorrect definitions and perceptions.

You can read about this in textbooks. The "Antenna Engineering Handbook" by Jasik in 3-13, 19-3, and in other sections uses correct definitions and descriptions.

Quad's and other Loops

We find the same efficiency misconceptions in articles about small loops and large quads. Authors sometimes assume, incorrectly, radiation resistance changes in a favorable proportion to loss resistance as we make the feed impedance increase. What we really are doing is placing the feedpoint in series with a smaller portion of NET current causing radiation.

With a large full-size quad element the pattern under some conditions will slightly change, but efficiency remains basically the same. With a small loop antenna, losses can actually increase with more turns!

Terminated Folded Elements

Another abuse of radiation resistance is found in terminated antennas. Some manufacturers and authors claim a resistance can be inserted in series with one leg of a folded monopole or folded dipole, and the other leg fed. The usual arm-waving claim is the antenna isn't really resistor loaded, and that efficiency is very good because radiation resistance is high.

That claim is absolute nonsense!

A large terminated Rhombic is well-known to have poor efficiency. Rhombic gain is actually low compared to other antennas having the same sin/sin x antenna pattern, because Rhombic efficiency is generally less than 50%. At least half of the power is consumed in termination and ground losses below the antenna. The actual gain may be reasonably high compared to a dipole, but not to other efficient antennas with the same half-power beam width.

The typical manufacturing buzz-word is that terminated monopole and dipole antennas are "traveling wave antennas" and by some magic (that even large terminated Rhombic antennas can't achieve) have broad bandwidth and high efficiency.

A Rhombic focuses energy (that is not transformed into heat) into a narrow beam that has considerable gain, but if it sprayed the radiation around in a non-focused pattern, a regular dipole would win hands down. Throw a resistor on that dipole to smooth SWR variations, or on a vertical, and efficiency suffers.

I listened to a station on 75 meters 600 miles away testing a Sommer T-25 vertical. He was 30 over nine using a dipole, and dropped to S6 with the vertical. The European he was working reported a similar change. By removing the termination resistor and base-loading the same vertical, a local Ham gained almost 25dB on 80 meters!

When we abuse or misuse radiation resistance, we can invent all kinds of magical antennas. We can have CFA's, E-H antennas, terminated dipoles, small magnetic loops, and verticals with all sorts of magical claims. Few, if any, of the claims are ever correct. Any time we see a claim that efficiency changes a ***large amount*** because of feed method change, it should be a red flag.

Increasing Radiation Resistance

Radiation resistance, at least under the useful IRE definition, can be defined by the following formula:

$$Rrad := 160 \cdot \pi^2 \cdot \left(\frac{He}{\lambda}\right)^2$$

which would translate to:

$$Rrad := 1580 \cdot \left(\frac{He}{\lambda}\right)^2$$

Where *He* is the effective height center of accelerating charges that cause radiation. In other words,

He is the effective height, expressed in fractions of a wavelength, of the distributed common-mode current in the structure.

(Common-mode current is the vector sum of all currents, or the effective in-phase current at any point, or the current we would measure if we placed a giant clamp-on current probe around ALL of the conductors at that any given height.)

He and λ must both be in the same units, either given as degrees or decimal fractions of a wavelength.

As an example, a uniform current single conductor antenna has an actual physical height of 15.19 feet on 1.8 MHz, where one wavelength is 546.67 feet, the effective height is:

15.19/546.67= 0.0278 wl

Since charges are distributed evenly throughout the structure the full height is used. The effective height is .027 wl, the same as the physical height. The height in electrical degrees is .0278 * 360 = 10 degrees

We have a radiation resistance of:

$He := .0278 \qquad \lambda := 1$

$$Rrad := 160 \cdot \pi^2 \cdot \left(\frac{He}{\lambda}\right)^2$$

$$Rrad = 1.22$$

We can express this graphically in a chart, such as one found in the Antenna Engineering Handbook by Jasik:

Finding 10-degree height on the graph above, and following that line until we reach the crossing for

unity current ratio, we see the ~1.27 ohm radiation resistance is in agreement.

Notice that the number of vertical conductors does NOT enter into the equation! This is the absolute maximum possible radiation resistance we can obtain for a given radiator height.

Non-uniform Current

Radiation resistance is purely a function of the effective current distribution and height of the radiator, and is limited by height (spatial length)! Current throughout the antenna will not remain uniform if we reduce the size of the flat-top or hat.

Current will become zero at the very top with no hat, and 100% base loading. In this case, with no change in height, radiation resistance will be approximately 1/4th the value of the uniform current example. The result is exactly like a 50% reduction in effective element height.

If we follow the 10-degree line to the intersection point with 0 top current, we find radiation resistance to be around .32 ohms. 1.27 ohms, the radiation resistance for uniform current, becomes 1.27/4 or .3175 ohms.

If we stay on the uniform current line, we find that .3175 ohms would be the radiation resistance of a 5-degree monopole with uniform current.

Efficiency

It often helps to look at the extremes, so we can get a feel for the effect of changes.

Let's look at the poor ground extreme and assume we have system losses, normalized to the current maximum, that are many times the radiation resistance. This would be the case for a short 160 or 80 meter mobile antenna.

In such a system radiation resistance would dominate any change that would affect efficiency. Current distribution would mean everything to efficiency.

Assume we have a base loading coil, either good or poor, and a thin mobile whip above the coil. Efficiency would increase by a factor of approximately four times by installing a capacitance hat with several times the distributed capacitance of antenna conductors below the hat.

Moving the coil would have little or no effect on efficiency.

A six-foot antenna with a large hat would be electrically equal to a 12-foot antenna without a hat.

This is why very poor inductors used on antennas in mobile shootouts, with large hats, equal or beat very large high-Q coils in similar height antennas that do not have large hats. One case in mind was a Hamstick lash-up in a mobile. The Hamstick, a notoriously poor efficiency antenna for 75-meters, soundly trounced Bugcatcher antennas when a large hat was added to the Hamstick.

Moving the coil up on the antenna has the effect of making current below the coil uniform, but without a hat current above the coil is a triangular taper that reaches zero at the element tip. The effective height of the area above the coil is 50% of actual height.

If we add a large hat at the bottom of the whip, current in the whip is actually reduced! At the same time, we change nothing below the coil. The effect of adding a large hat below the whip is to reduce the effective height of the antenna, when considered as a percentage of physical height. Radiation resistance and efficiency is generally reduced by adding a hat just above a coil, even if the hat allows us to use a smaller coil!

Adding a large hat at either end of a coil also reduces coil Q, since a large portion of the hat capacitance directly shunts the inductor.

Conclusion

We can reach the following conclusions:

- Radiation resistance, or at least the useful definition of radiation resistance, is limited by spatial area (or height in the case of a vertical) any antenna occupies.
- Radiation resistance is maximized by making current as large as possible over the entire spatial area of the antenna.
- Surrounding objects generally reduce radiation resistance and efficiency, even when they are NOT resonant, because they reduce effective height! This includes dielectrics that increase capacitance of the antenna to ground, since any increase in capacitance appearing well below the top of an antenna reduces effective height.
- Radiation comes from charge acceleration, nothing else. The longer the linear spatial distance we move charges in, the fewer charges we need to move at any point for the same amount of EM radiation. This is just another way of saying radiation resistance is higher in physically longer structures, especially when they carry uniform current.
- Any antenna (including Linear Loading, Helical Loading, Folded Monopoles, Fractal shapes, CFA, E-H, and so on) claiming to increase radiation resistance beyond the limits outline above is based on misunderstandings or untrue distortions of basic antenna principles.

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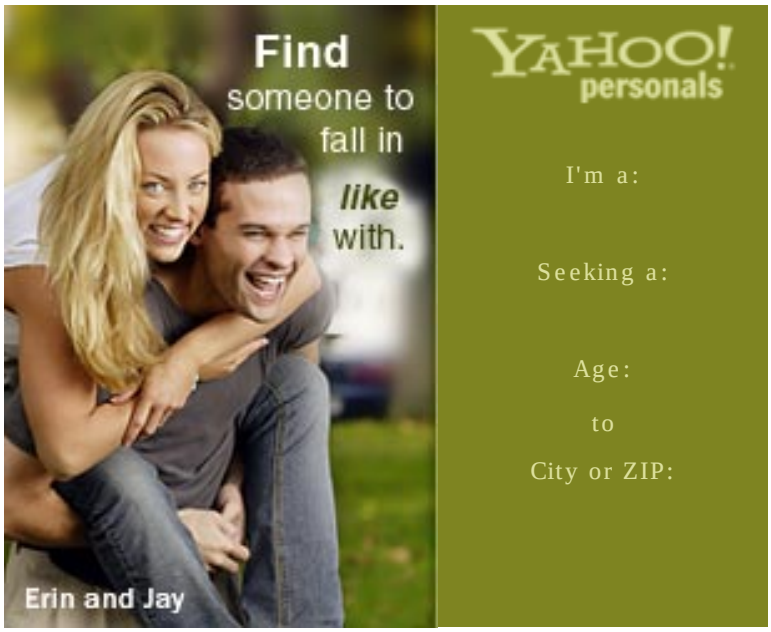

Slingshot method for hanging SkyWires.

Guys,

In my youth (19 years old), fresh out of Marine Corps Boot Camp, I would climb to the tops of trees like a monkey, while friends on the ground would yell "that's high enough!!" to put up Owl Nesting boxes, etc.. What they didn't know was it was much easier that most of the stuff the Corps had me doing.

That was then and this is now. Now Walter no longer leaves the ground unless I have a boarding pass, and a coach ticket.

After reading a few articles I decided to try the slingshot

method of
hanging a wire antenna.

Its Pretty easy.

You need the following items.

A fishing pole, a dozen 1 oz. Lead egg sinkers, a good Crossman or
Wrist-Rocket slingshot, a large spool of Trout Line, Large Nails,
a hammer, and a Fishing pole holder, if you can't find someone
to
help.

I've seen ads in radio magazines for a slingshot-fishing reel combo,
for $100. If you have $100 to blow, take you wife out to a fancy
restaurant. I believe it's a much better investment that this.
The
slingshot, and a fishing rod and reel will set you back about $30..

I use trout line because it is very strong, and given its to be used
in water, its waterproof, and seems to hold up very well, and is
cheap. But if someone knows of another rope that works better, please
let me know.

You can find all of this at your local bait and tackle store. I pay
about 20 cents for each egg sinker, I use egg sinkers because they
are about as aerodynamic as fishing weights get, and because the
shape doesn't allow them to get hung up in the tree branches as
much
as others do, but it still happens.

I have heard of people using a Bow and Arrow. I can see some
advantage to this, but I think the risks outweigh the benefits. With

a 1 oz. Egg sinker, you have to be careful it doesn't come down on a parked car's windshield, or dent the hood. With an Arrow, it could put a hole in someone's roof, or worse, if you neighbor or his kids are outside, they could end up impaled and looking like Gregory Peck in the last scene of the movie "The Omen". We don't want that…

When shooting the egg sinker, you want to be on the "Antenna side" of the tree.

Take the pole holder, and place it in the ground at an angle, facing the target tree. Put the fishing pole in the holder.

I have read that if you paint the egg sinkers blaze orange, you can see them better when they come down. I haven't tried this, but it makes sense. Having a florescent fishing line in your fishing reel helps in locating the downed egg sinker, but is not necessary.

If you are right handed, taking the slingshot in the left hand, and stand with the fishing pole to the right of you. If left handed, do the opposite. Place the egg sinker in the sling with the fishing line coming out of the top, and clear from the rubber bands.

Make sure not to stand too close to the tree, you want the egg sinker to clear the tree, and fall on the far side, if you are too close, you will be shooting at too high an angle and the egg sinker won't fall on the far side, but on the shooting side.

Shoot the egg sinker and pay attention to the direction of

where you
shoot, and see if you can judge where it should have landed before
you walk over to the far side of the tree.

Then go find it. You may find that you need to take the fishing pole
and "play" with the line a bit to get the egg sinker to fall down to
ground level from the higher branches.

If it gets hung up, you will just have to reel it back and start
over. I have about a 50/50 success rate.

Once I find the egg sinker, I cut it off, then tie the trout line to
it and go back to the fishing pole and reel the trout line back over
the top of the tree.

When the trout line reaches the pole, cut the line, and tie your dog
bone insulator on to it.

Go back and cut the trout line from the trout line spool, and take
two large nails and hammer each of them about half way into the trunk
of the tree. The nails should be about a foot or so apart and angled
away from each other forming a cleat.

I like to hammer the nails high enough that some kids don't come
along and decide to undo my work. If this is a real concern, you may
consider using a ladder and nail them higher up on the trunk of the
tree. Plus, the higher up, generally the less visible it is too.

Tie the trout line around one nail.

Repeat these steps on all the trees in your SkyWire system.

Once you run your antenna wire through the dog bone insulator

you can
then begin to raise the antenna by taking the slack out of the trout
line. Just begin wrapping the trout line around the both nails in a
loop until the antenna reaches the desired height.

That's about it.

I am currently working on a fuse system that uses a bungee cord, and
counter-weight, so that when bad weather hits, and high winds put too
much pressure on the SkyWire, that the "fuse" will break causing
about 20 feet of extra trout line to be unleashed, and letting the
SkyWire fall limp, but not letting it break. This would allow you to
easily recover the extra trout line, and put a new "fuse" back into
the system after the storm passes.

Once I have the dynamics of this worked out, I will post it.
(sometime this fall, I hope)..

Thanks,

Walter
http://groups.yahoo.com/group/SkyWires/

	Replies	Name/Email	Yahoo! ID	Date
38	Re: Slingshot method for hanging SkyWires.	rickfrazier96720	rickfrazier96720	Tue 9/9/2003
40	Re: Slingshot method for hanging SkyWires.	Clinton	ve7knl	Wed 9/10/2003

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Traps

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This article is from a talk I gave about antenna traps. It contains measurements of traps, performance of trap antennas using models of traps, and ideas on how to make trap antennas more efficient.

Try taking this Trap-Q test! Be honest.

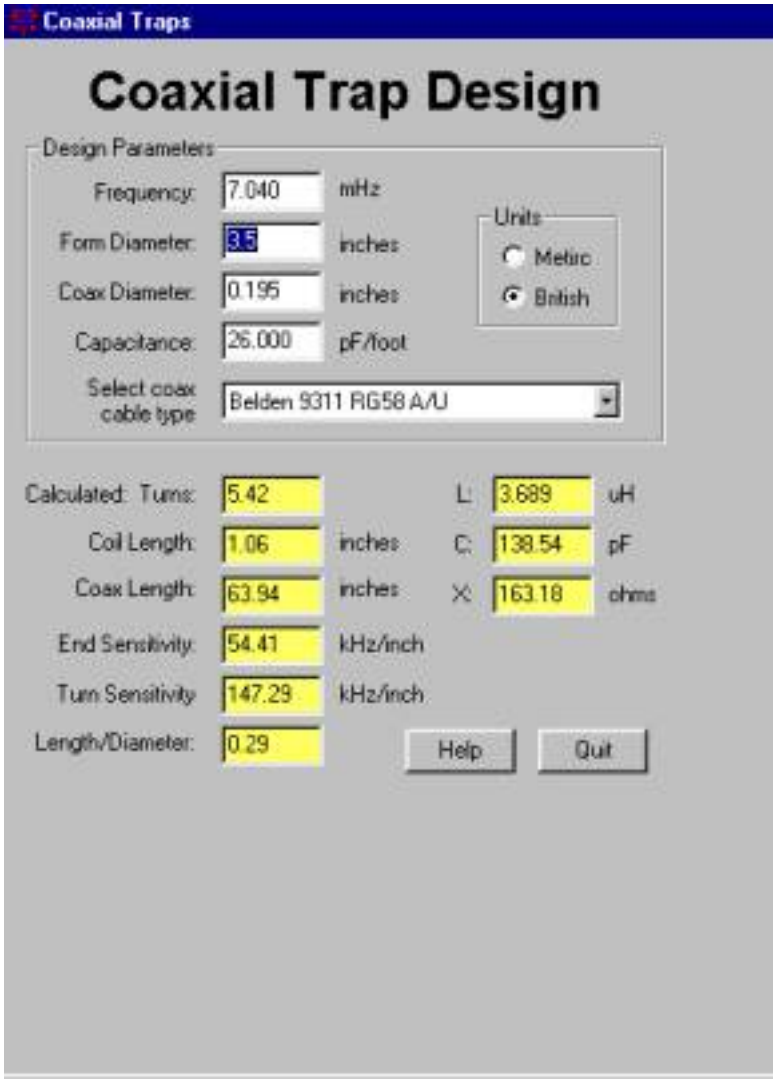
1.) Is it best to make the trap resonant close to the desired operating frequency?

2.) Does bandwidth decrease with increasing trap Q?

3.) Do traps create noticeable loss, perhaps one dB per trap typically?

4.) Does higher trap operating Q always mean lower loss?

Coaxial Trap Designer by VE6YP (Tony Fields)



This is a good program to get you in the ballpark with a trap design. It was available as freeware. (Unfortunately coaxial traps are relatively lossy on the trapped frequency compared to other types.)

The software is available at **http://members.shaw.ca./ve6yp/index.html**

7.04 MHz 3.5 inch diameter form RG58/U into the VE6YP program yields calculated values of:

Calculated	Actual Measurement
L= 3.689µH	3.116µH
C= 138.5 pF	164 pF
64 inches	59 inches

Using the program TLA by N6BV (from ARRL), we would estimate capacitance of a 59" RG-58/U cable as:

R .22 –j143.61 or about 157 pF (Q=650)

Measuring a real-world stub, capacitance was 164pF (Q=590).

While that Q seems high, remember a typical transmitting-type air-variable capacitor has a Q of several thousand!

Coaxial Trap Articles and Programs use capacitance/ft multiplied times length….

26 pF * 4.917 feet = 127.84 pF in trap program

C164 pF measured. This error, 36pF low from 164pF, occurs because the transmission line making up the "coaxial capacitor" is not actually treated as a transmission line in the modeling program.

Fortunately the error is in a useful direction, because we can shorten the cable! Coaxial capacitors are really open stubs, and should be treated that way once they are more than a few degrees long.

CONCLUSION: The difference between TLA and an actual measurement was around 4%. This is very close, but the result has significant difference from the coaxial trap program since it only considers pF per foot as the capacitance. A longer cable (in fraction of a wavelength) results in greater error by using pF per foot. The error comes because a coaxial cable capacitor is really a stub, NOT a pure capacitor!!

Trap Measurements (at resonance)

Type	F MHz	R parallel	X
Coax RG-58	7.034	17,800	0
UT-141-75 semi-rigid	7.045	45,330	0
100pF 7.5kV & #12 wire	7.040	99,850	0
60pF 15 kV & #10 wire	7.040	250,000	0
60 pF vac & Copper tubing	7.040	300,000	0
Coax RG-58	3.700	23,200	0
Coaxial with fixed mica capacitor	7.040	21,660	0

Highest R parallel equivalent is best!! Lower Rp means more loss.

Trap Measurement summary:

- **Coaxial trap poorest**
- **Once #10AWG wire is used, not much improvement**
- **Space-wound bare wire makes best inductor**
- **Transmitting-type capacitors noticeably better than capacitors made from coax**

10 Meter (Tribander) Traps

Type	Freq	R parallel	X
Coax RG-58	29.00	13,800	0
Mosley TA-33	30.64	43,100	0
Mosley Pro-57	27.46	66,080	0

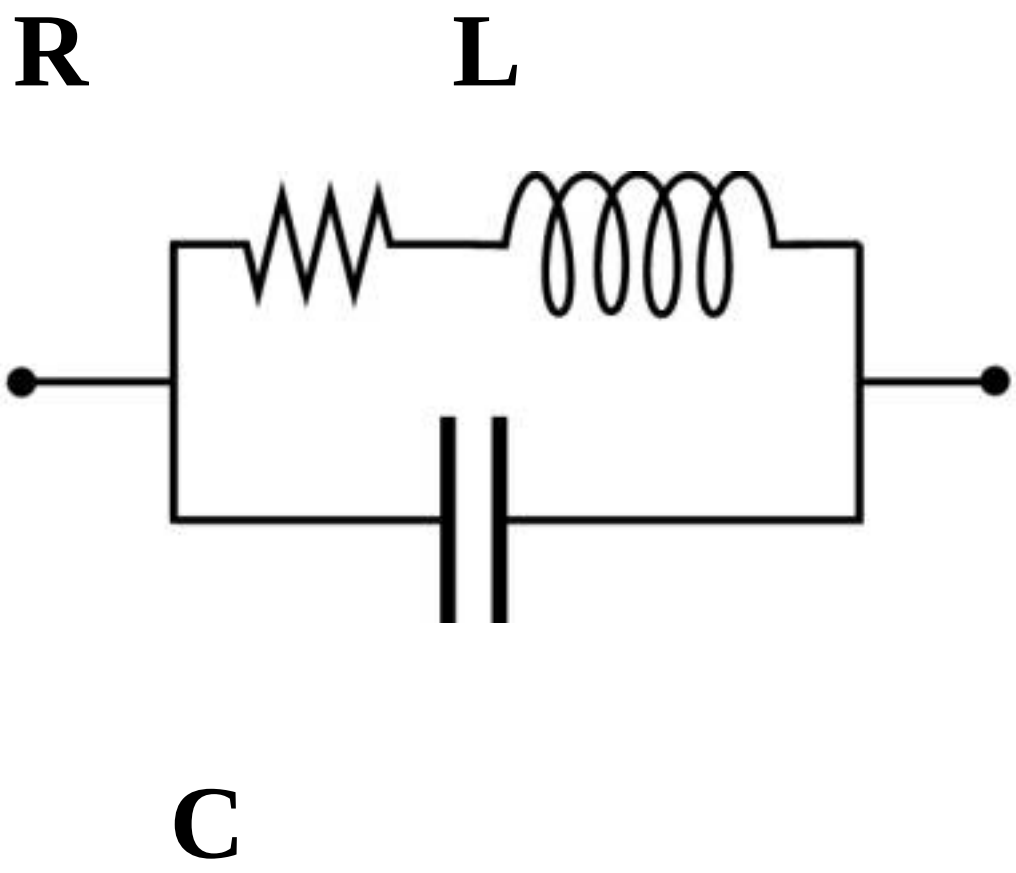
Cushcraft A3	**28.78**	**110,000**	**0**
HyGain TH-3	**29.67**	**140,200**	**0**

Traps are not all that bad when you plug them into models.

15 Meter (Tribander) Traps

Type	Freq	R parallel	X
Coax RG-58	**21.00**	**13,980**	**0**
Cushcraft A3	**21.43**	**76,270**	**0**
Mosley TA-33	**21.68**	**79,000**	**0**
HyGain TH-6	**22.23**	**142,000**	**0**

Trap Model



Measured Values Coax 7 MHz Trap

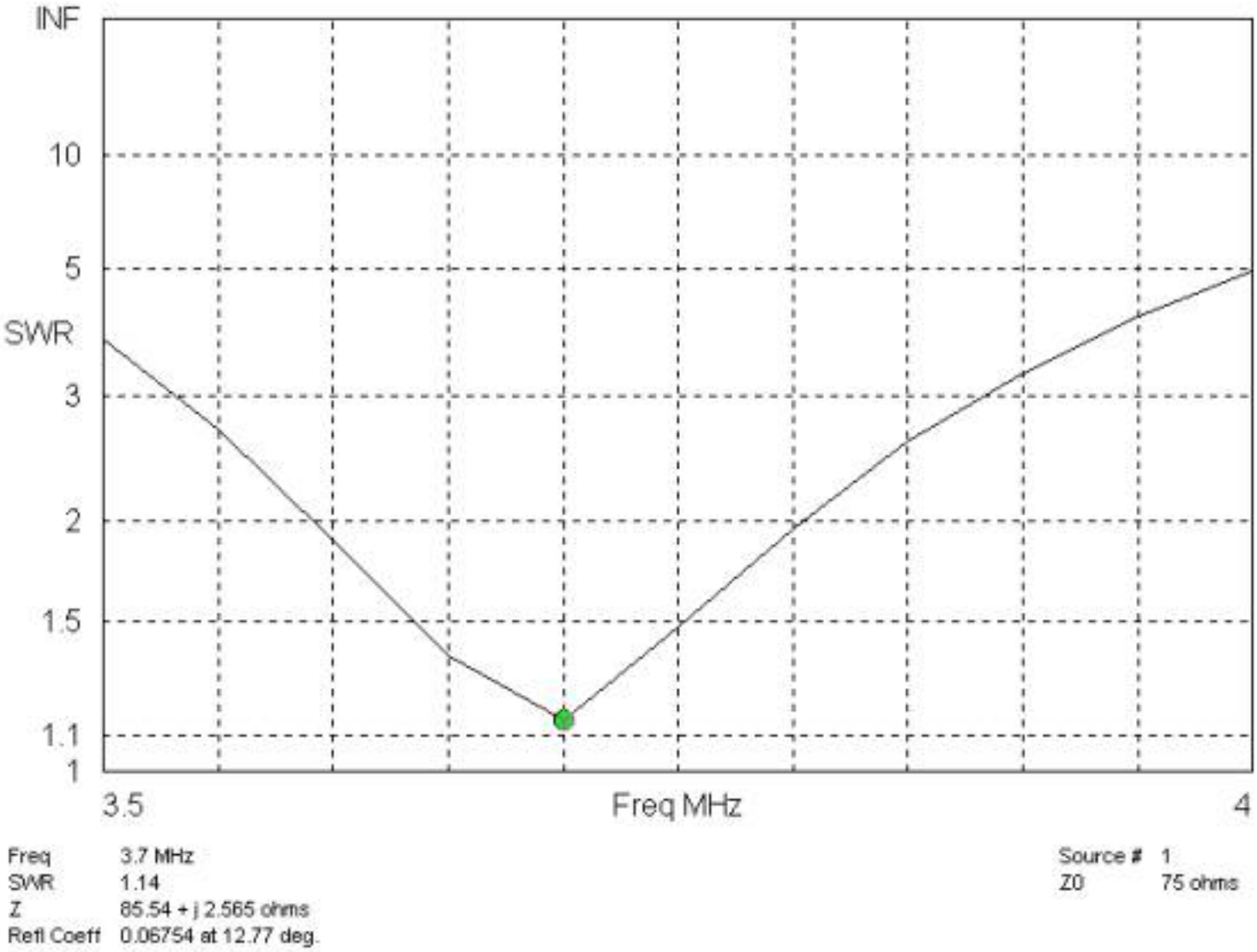
Freq	Imp	R	Xc	Q	L uH	C pF
7.04	17,800 j0	1.03	138	134	3.114	164

3.7	1.1 j 97	.6	283	88	3.114	152

Measured Values L/C 7 MHz Trap

Freq	Imp	R	Xc	Q	L uH	C pF
7.04	99,850 j0	.36	215	465	4.92	105
3.7	0.53 j 156	.25	409	294	4.92	105

SWR Bandwidth



7 MHz RG-58 TRAP

80 m 75 ohm VSWR

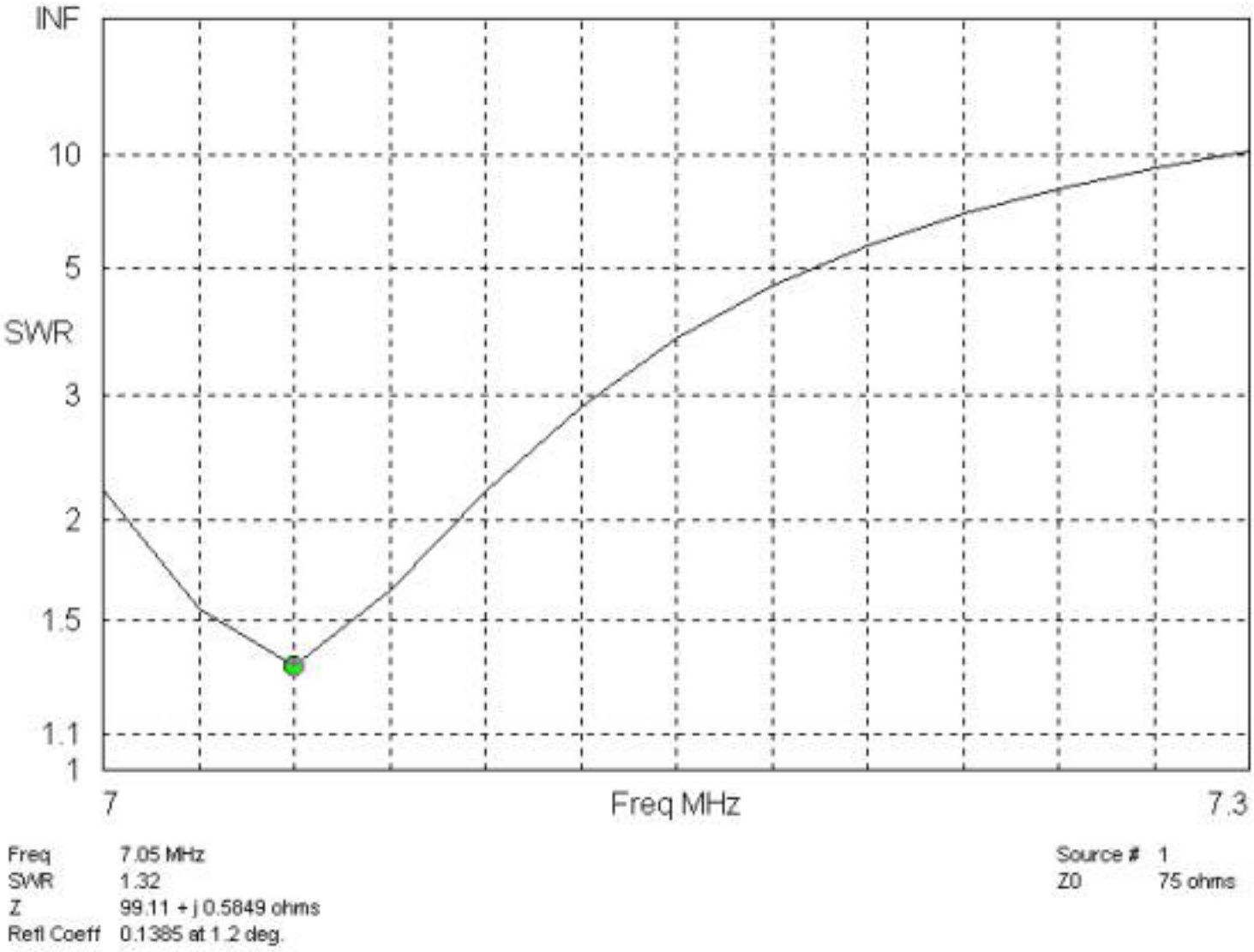
EZNEC #12AWG dipole

Coax trap 80m 2:1 VSWR ~210 kHz

Total trap loss = 0.05 dB

RG-58 TRAP, 75 ohm VSWR, 40 METERS

VSWR BW



Coax trap 40 meter 2:1 VSWR ~ 80 kHz

Total coaxial trap loss at resonance on 40m= 1.6 dB

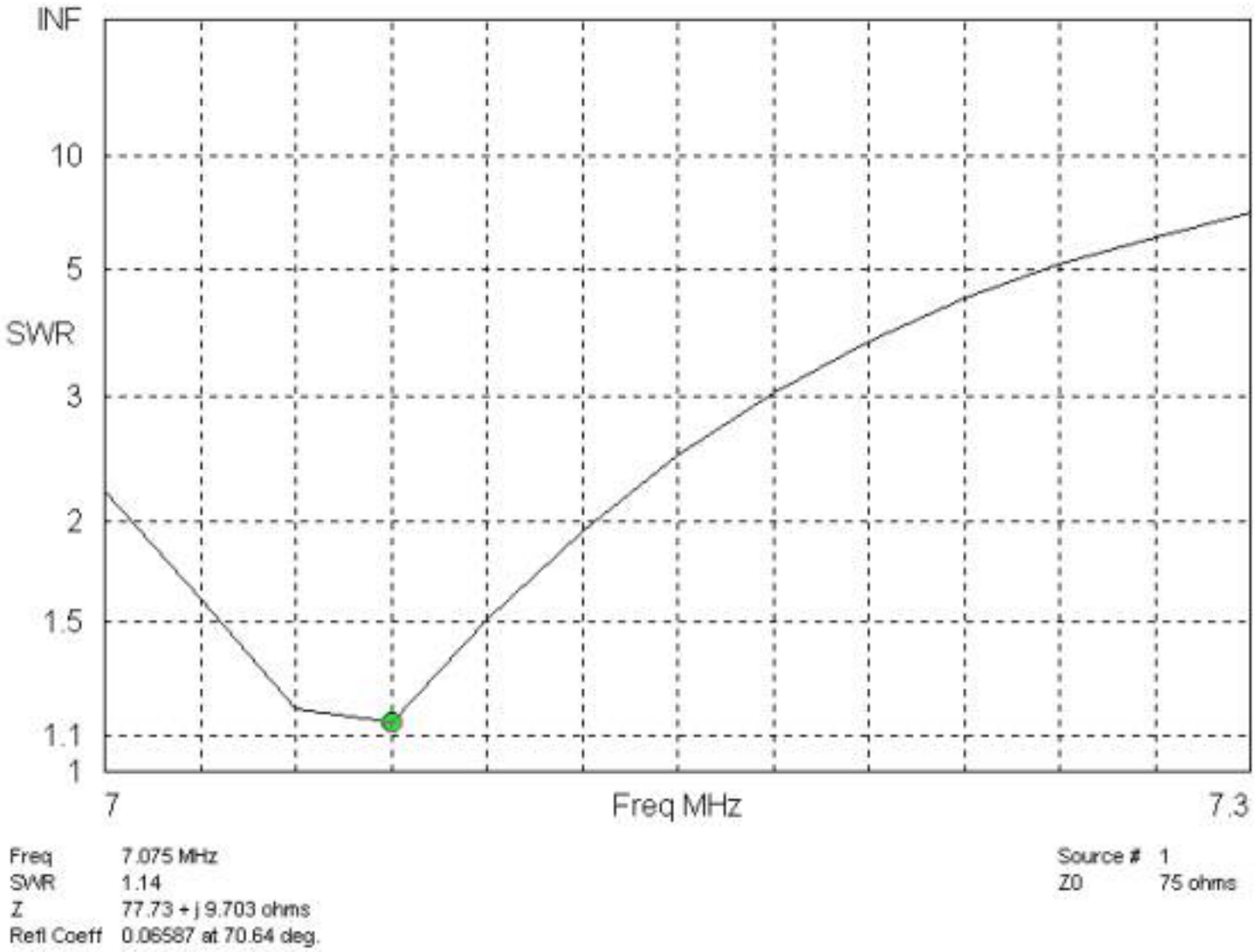
Total coaxial trap loss 100kHz off-resonance (at 7.15 MHz)= 1.06 dB

Note that loss is maximum at trap resonance!!!

Never make a trap resonant on the desired operating frequency!!

W2LH ARRL Handbook Trap Design

100pF #12awg Miniductor trap

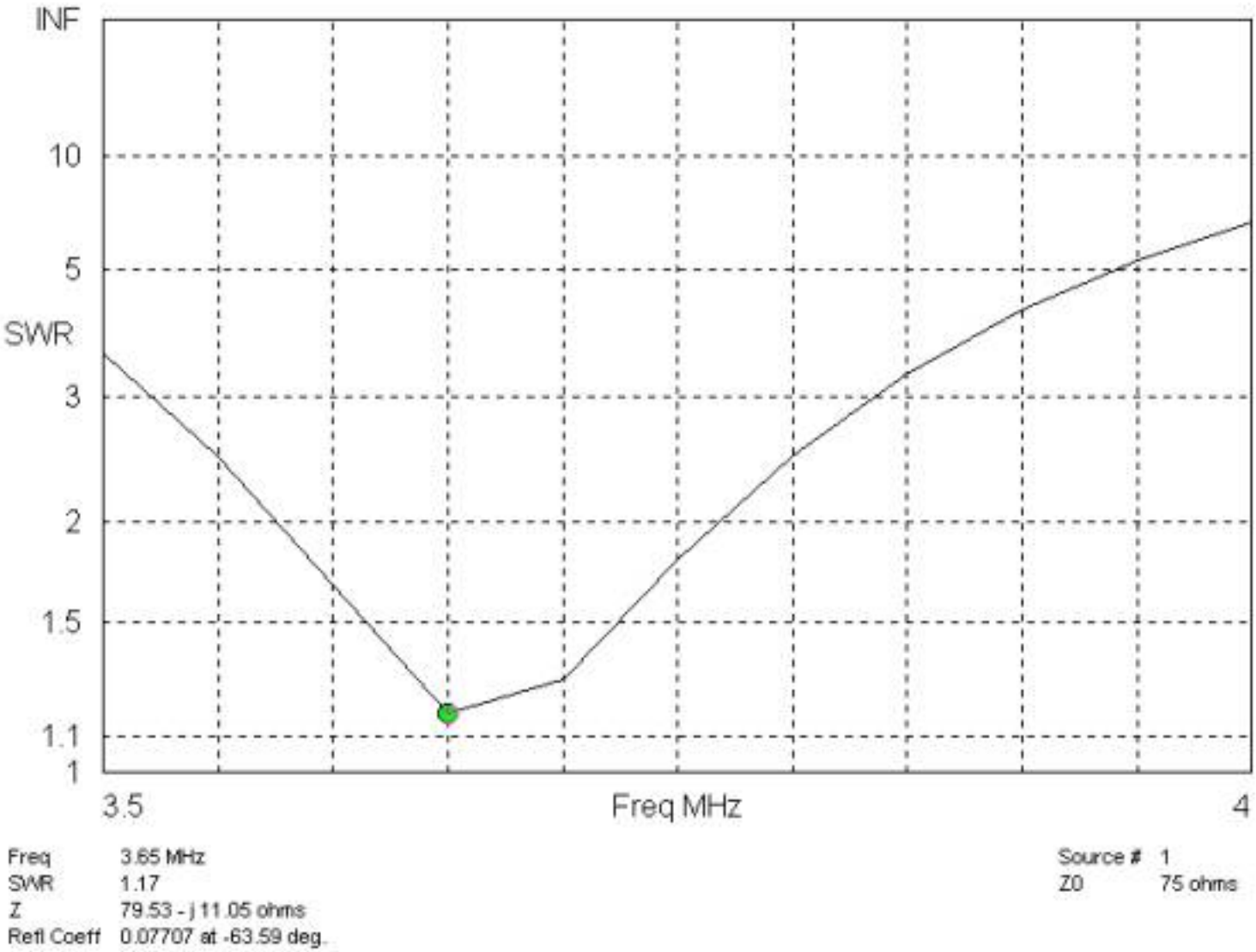


40m 2:1 VSWR ~120 kHz

Total loss = 0.24 dB

W2LH ARRL HANDBOOK TRAP 80m

VSWR

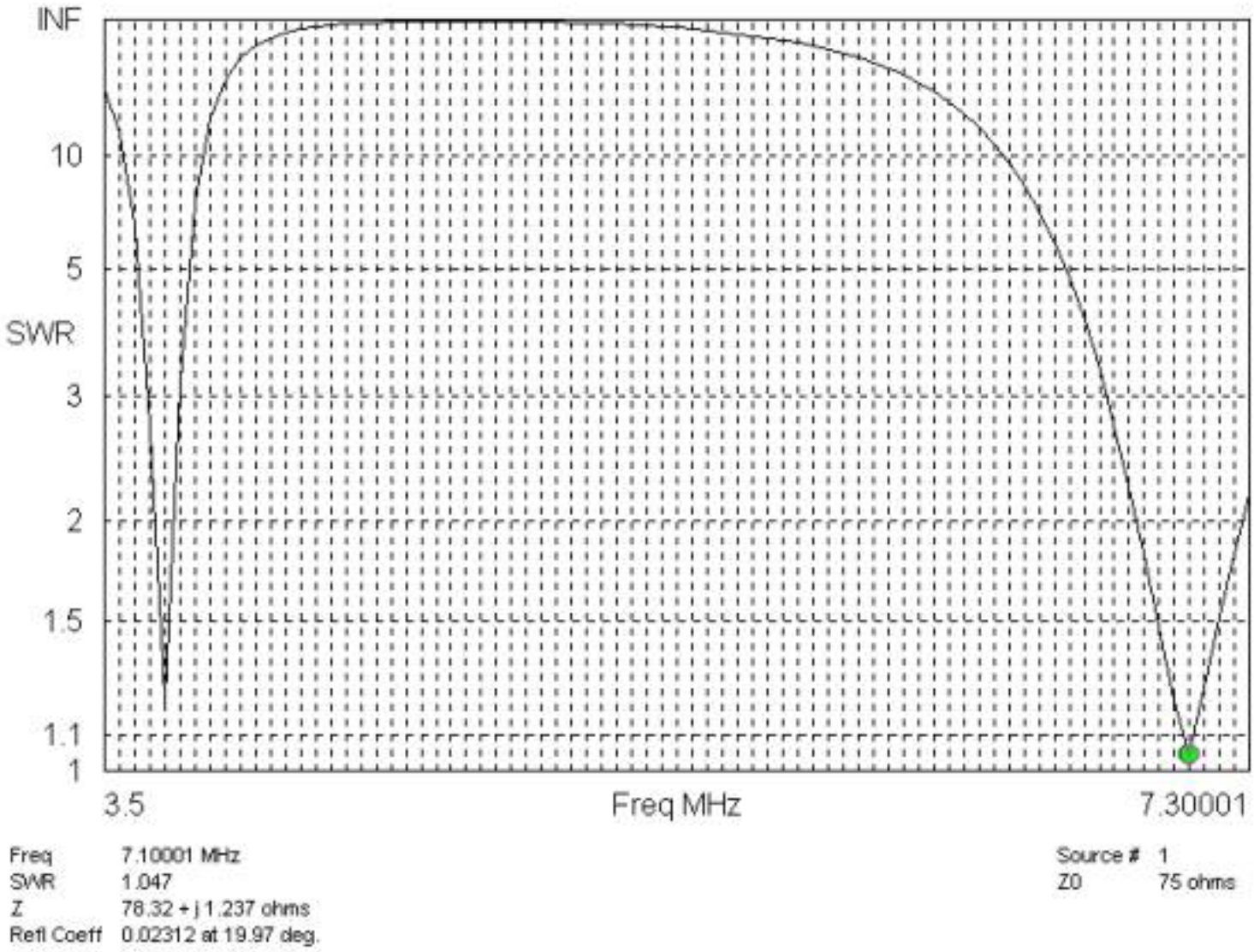


80m 2:1 VSWR BW ~ 200 kHz

Total trap loss = 0.026 dB

What happens if trap is not in band?

VSWR Bandwidth of 6.51MHz trap in 80/40 dipole



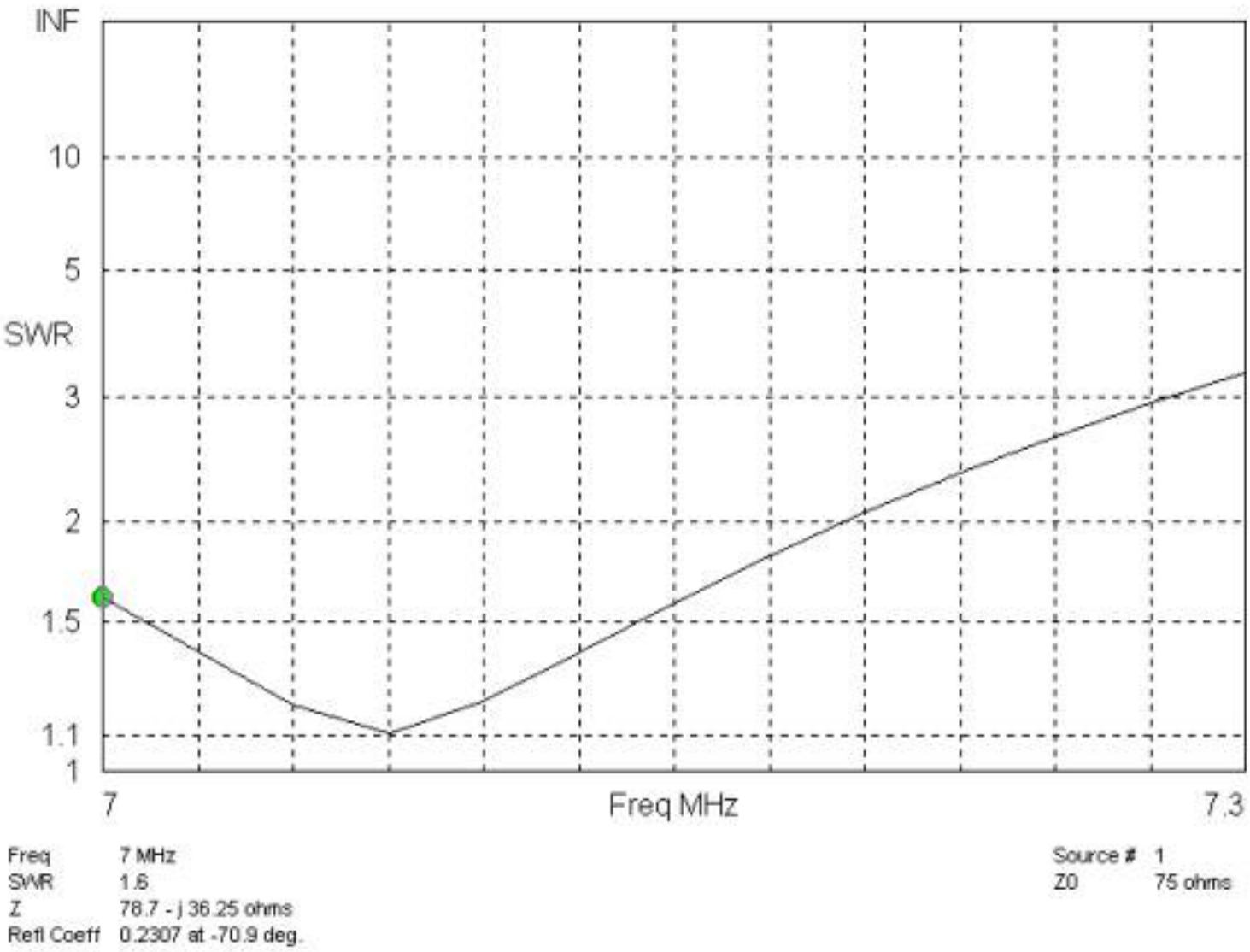
Trap at 6.51 MHz Q=130

Loss at 7.15 MHz = 0.314 dB

Loss at 3.7 MHz = 0.324 dB

This is a 104-foot long antenna, with very poor Q traps, and loss is less than .4dB! The reason loss is low is we have moved the trap slightly out-of-band.

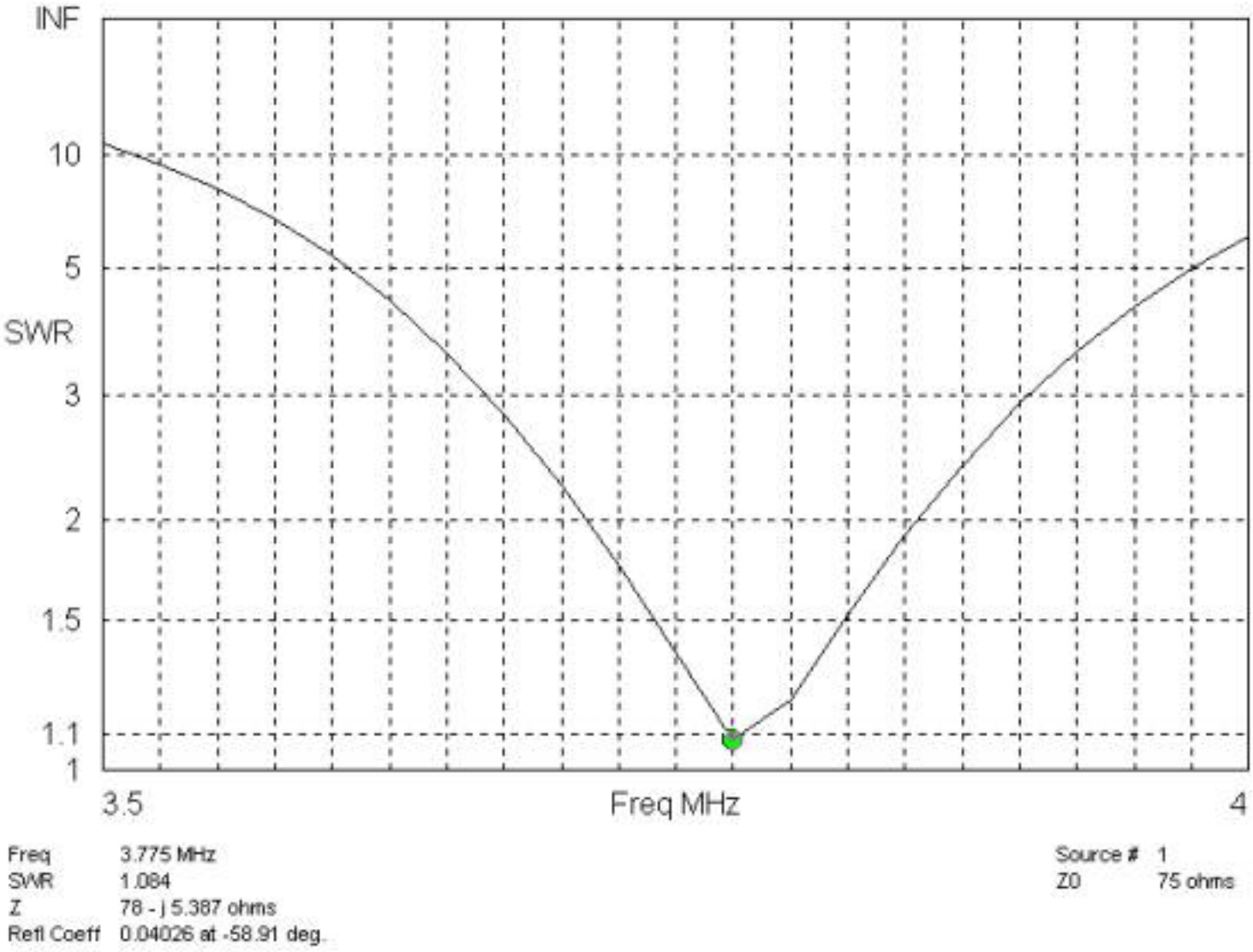
6.15 MHz Q=130 TRAP 40m VSWR



7 MHz 2:1 VSWR BANDWIDTH ~200kHz

Trap Q at resonance = 130 7 MHz loss ~ .3 dB

6.15 MHz Q=130 TRAP 80m VSWR



80M 2:1 VSWR BW ~130 kHz

Loss at 3.7 MHz = 0.324 dB

1.) Is it best to make the trap resonant close to the desired operating frequency?

NO! Loss is highest when the trap is resonant at the operating frequency!

2.) Does bandwidth decrease with increasing trap Q?

NO! Bandwidth is a function of many variables, trap Q actually has one of the smallest influences on BW.

3.) Do traps create noticeable loss, perhaps one dB per trap typically?

NO! Even the worse traps (coaxial traps) in the worse possible condition of operation are only 1.6dB loss for BOTH traps!

4.) Does higher trap operating Q always mean lower loss?

NO! Loss depends on many factors, including trap resonant frequency.

Conclusions:

- Trap loss has been greatly exaggerated by advertising hype
- Traps should not be resonant at the actual planned operating frequency
- Coaxial traps are more lossy than articles conclude
- Coaxial stubs used as capacitors can not be calculated using pF/ft unless the stub is a very small fraction of a wavelength long (less than a few electrical degrees)
- Coaxial stubs have low Q (are relatively lossy) compared to normal lumped components.

Aerials and Trees

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Click ***HERE*** for the Amazing **Online Ham Radio Flea Market**

6.7m fibreglass poles for sale click ***HERE***

From time to time we all need to get wires or lines over trees to support aerials. With a small tree of say 5-7 metres high it is quite easy just to throw a weighted line over the top. But what if the tree is larger?

Over the years I have tried quite a few ways but the best by far has been the use of a catapult - suggested to me by Dave G3YXM who also uses a bow and arrow! Catapults can be bought from fishing shops and are used to throw ground bait. Look for a high power model and for maximum accuracy and power, choose one with an arm brace (see picture below). But don't leave the fishing shop just yet as you will need some more things. First get a cheap casting reel. They are often available for about £5. This needs to be fastened on to the catapult. I used some cable ties. The casting reel ensures that the line flows freely. Laying the line on the ground generally fails as it gets caught or tangled. For line, you need to choose something thin and light - this line will not be used to support the aerial, it is just to get it over the tree. Use 12-15 pound breaking-strain monofilament. You will also need a weight to fire. I use a 2 oz sinker. I paint them orange to make them easier to see in trees! Tie the weight on to the line using the special knot that works with monofilament (called a blood knot I think). This knot is described in fishing books. I have also had success with a figure-of-eight knot.

Search G3CWI's site

Now you are all set. Safety must be your first consideration. Make sure that the area around the tree is clear of any people, animals, overhead electricity cables and anything else that could either endanger you or could be in danger from your efforts. Fire the weight aiming well clear of the top of the tree, an angle of between 45 degrees and 60 degrees to the horizontal seems best. If all is well the line should pay out cleanly but generally the weight will not reach the ground. Gently pull on the line several times, letting the weight drop further each time. Then go and see if you can find it. The paint helps here! If you fail to get it in the right place, you have two choices, either cut the line and let the weight drop or try to pull it back. If you do try to pull it back be very careful as the weight can easily fly out of the tree in your direction as you pull on the stretchy monofilament.

If all is well and the weight has reached the ground in the right spot, then cut the weight off the line and tie on a thicker line that will be used to support the aerial. I use 40 pound or 100 pound monofilament. Pull this line back through the tree, tie on the aerial, pull it up and you're on the air. I have used this method successfully with trees up to 90 feet tall (27 metres).

The G3CWI Catapult in action

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Beam Antenna Accessories
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Vertical Antenna Accessories
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Description	File Name	SizeKB
Beam Antennas		
Model 10/11, Butterfly beam antenna covering 10 or 11 meters produced from 1987 to Present	00246IZV	81
Model HF3B, Butterfly beam antenna covering 10, 15 and 20 meters produced from 1998 to Present	00761IZV	
Model HF3B/HF4B, Butterfly beam antenna covering 10,12, 15 and 20 meters produced from 1985 to 1985	00703IZV	575
Model HF4B REV A, Butterfly beam antenna covering 10,12, 15 and 20 meters produced from 1985 to 1986	00402IZV	566
Model HF4B REV B, Butterfly beam antenna covering 10,12, 15 and 20 meters produced from 1986 to 1987	00704IZV	1,032
Model HF5B, Butterfly beam antenna covering 10, 12,15, 17 and 20 meters produced from 1987 to 1988	00706IZV	766
Model HF5B REV A, Butterfly beam antenna covering 10,12, 15, 17 and 20 meters produced from 1988 to 1991	00810IZV	847
Model HF5B REV B, Butterfly beam antenna covering 10,12, 15, 17 and 20 meters produced from 1991 to 1995	00705IZV	1,013
Model HF5B REV C, Butterfly beam antenna covering 10,12, 15, 17 and 20 meters produced from 1995 to Present	00159IZV	123

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Model 18MCA, 17 meter kit for the HF3B or HF4B Butterfly beam produced from 1987 to Present	00370IZV	18
Model 24MCA-II, 12 meter kit for the HF3B Butterfly beam produced from 1985 to Present	00373IZV	16

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Model 2MCV-5, VHF collinear vertical covering 2 meters, 5 dB gain produced from 1984 to 1994	00244IZV	123
Model HF2V, HF Vertical antenna covering 40 and 80 meters produced from 1984 to Present	00156IZV	77
Model HF4V-II, HF Vertical antenna covering 10, 15, 20, and 40 meters produced from 1978 to 1981	00409IZV	188
Model HF4V-S, HF Vertical antenna covering 10, 15, 20, and 40 meters - 14 ft produced from 1978 to 1981	00811IZV	235
Model HF5V, HF Vertical antenna covering 10, 15, 20, 40 and 80 meters produced from 1974 to 1977	00812IZV	179
Model HF5V-II, HF Vertical antenna covering 10, 15, 20, 40 and 80 meters produced from 1977 to 1979	00813IZV	379
Model HF5V-III, HF Vertical antenna covering 10, 15, 20, 40 and 80 meters produced from 1979 to 1981	00464IZV	306
Model HF5V-S, HF Vertical antenna covering 10, 15, 20, 40 and 80 meters - 16 ft produced from 1978 to 1981	00463IZV	205
Model HF6V, HF Vertical antenna covering 10, 15, 20, 30, 40 and 80 meters produced from 1981 to Present	00366IZV	91
Model HF6V-X, HF Vertical antenna covering 10, 15, 20, 30, 40 and 80 meters produced from 1987 to 1997	00233IZV	132
Model HF9V, HF Vertical antenna covering 10, 12, 15, 17, 20, 30, 40 and 80 meters produced from 1997 to Present Model	00408IZV	115
Model HF9V-X, HF Vertical antenna covering 10, 15, 20, 30, 40 and 80 meters produced from 1989 to 1997	00233IZV 00234IZV 00235IZV	132 18 19

Model SC-3000, Collinear vertical scanner antenna covering 30-512 MHz produced from 1984 to 1995	00245IZV	68
Model SC-5000, Collinear vertical scanner antenna covering 30-905 MHz produced from 1985 to 1995	00819IZV	63

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Model A-18-24, 12 and 17 meter kit for vertical antennas except HF2V produced from 1983 to 1989	00820IZV	114
Model A-6, 6 meter kit for vertical antennas except HF2V produced from 1987 to Present	00235IZV	19
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Model TBR-160-HD, Top band resonator kit (160 meters) for all vertical antennas produced from 1981 to 1987	00822IZV	147
Model TBR-160-S, Top band resonator kit (160 meters) for all vertical antennas produced from 1987 to Present	00237IZV	26
Model TBR-160-S-NK, Neutralizing kit for the TBR-160-S produced from 1987 to Present	00238IZV	7
Model TLK, Top loading kit for the HF2V produced from1984 to Present	00242IZV	9

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Santa Clara County, California
ARES/RACES

EMERGENCY OPERATIONS & YOUR HT

Before I begin, I would like to give credit to C. Edward Harris, KE4SKY, AEC for Fairfax County ARES in Virginia for his article on "Getting the most from your Hand Held Transceiver." Much of this script has been taken from that article. Some minor editorial changes have been made to make it more readable over the air.

<PAUSE for the Repeater to reset >

HT Antennas

It shouldn't come as a surprise to anyone that when limited to "barefoot" operation with a "rubber duck" on simplex, HTs are not adequate as a primary rig for emergency communications. The National Institute of Standards and Technology (NIST) ran some tests (back when they were still the NBS) on Public Safety high band and amateur 2-meter antennas. They found that a "rubber duck" has -5db gain compared to a quarter wave antenna held at shoulder height. In terms of effective radiated power (ERP), a 5w HT with rubber duck antenna, held at shoulder height would actually radiate 1.5 watts. Placing the HT on your belt attenuates the signal another 20db, reducing ERP to only 15 milliwatts! UHF results weren't found to be much better.

<PAUSE for the Repeater to reset >

A simple and inexpensive improvement that can be made to the "rubber duck" is the addition of what is called a "tiger tail". You can make one of these using a quarter-wavelength (19-1/4" for 2 meters) piece of #14 through #20 stranded wire, crimped and soldered to a battery clip. Reinforce the soldered connection with heat shrink tubing or tape to resist flex. Clamped to the outer collar of the BNC connector on your HT antenna, it acts as a counterpoise so that RF from the HT doesn't couple with your body. A "tiger tail" is directional and can be used to change both radiation angle and direction. It gives best simplex performance when pointed in the general direction of the station you are trying to "hit".

<PAUSE for the Repeater to reset >

Almost any antenna works better than the "rubber duck" that comes with an HT. A

flexible ¼ wave or telescoping half-wave antenna are both improvements. A ¼-wave used at shoulder height with a counterpoise has "unity" gain, which is a 5 db improvement over a rubber duck, because most of the signal is radiated. Using an HT at 5 watts with a ¼ wave mag mount on an improvised ground plane, or telescoping half wave with a "tiger tail" improves simplex readability even further.

<PAUSE for the Repeater to reset >

In marginal locations, a telescoping half-wave is a good performer. A half wave used without a ground plane has the same unity gain as a ¼ wave when used with a ground plane. Adding an effective ground plane or counterpoise to a half- wave produces roughly 2 db of gain. A telescoping half wave can also be attached to a coax jumper and pulled into a tree, dangled out a window, attached to a window pane with suction cups or used with a window clip door mount.

<PAUSE for the Repeater to reset >

Telescoping antennas work best when operating stationary or in the open, avoiding side impacts or rough handling. Extend and retract the radiating elements very carefully. If you note any wobbling or looseness, replace the telescoping radiator, if possible, or replace the entire antenna. Keep a close watch on your HT's connector also. It can become loose after extended use of a telescoping antenna.

<PAUSE for the Repeater to reset >

Flexible antennas are safer when working in close quarters around people and are more durable when walking through dense vegetation during search and rescue operations. They are a good choice for dual-band transceivers, but are usually optimized for one band and merely "acceptable" on the other. Most approximate a ¼ wave on 2 meters and a 5/8 wave on 70 cm. How efficient a particular antenna is can be determined only by controlled testing.

<PAUSE for the Repeater to reset >

If you want to buy one emergency HT antenna, without risk or experimentation, the telescoping half-wave, flexible 5/8 wave or quarter wave mag mount all offer the best "bang for the buck" in my opinion. A telescoping half-wave boosts practical simplex range of a 5 watt, 2-meter HT from several miles with a rubber duck to many miles over suburban terrain. Adding a tiger tail further extends readable simplex range under the same conditions.

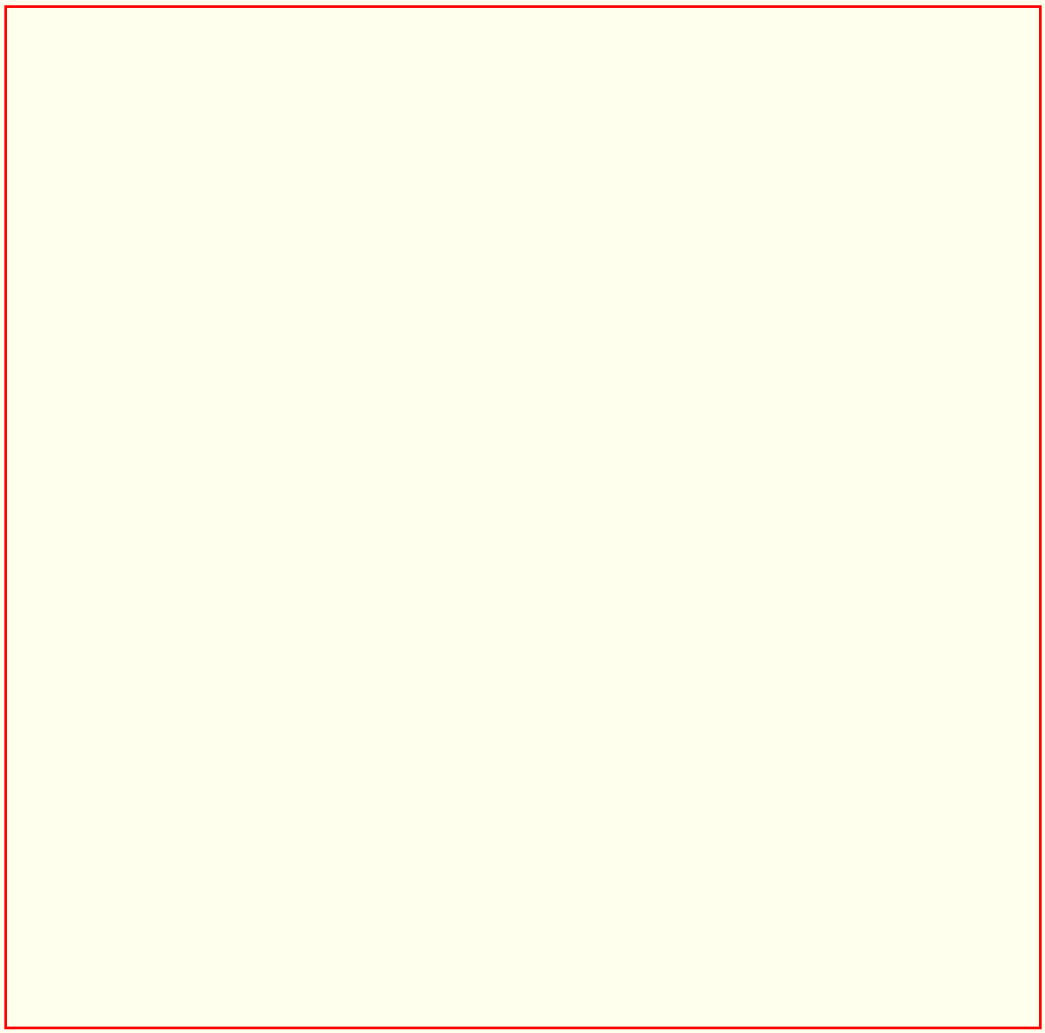
<PAUSE for the Repeater to reset >

Whatever antenna you choose, try to find one that is rated for at least 25W so that it can also serve as an emergency antenna for the HT with a power amplifier at medium power. A ¼ wave mag mount connected to a power amplifier works best on a car, but a suitable improvised ground plane can usually be found around the home or office. A metal filing cabinet, rain gutter, refrigerator, balcony railing or other large metal object may work just as well. If all else fails, place aluminum foil over a large piece of cardboard .

<PAUSE for the Repeater to reset >

A good possibility for Fixed position Emergency Operations is the so-called Roll Up J-pole . It's made from 300 Ohm TV Twin-lead and should give you several dB of gain over a rubber duck. I've used mine for the past year and one-half as a base antenna and have been quite happy with it. It is well worth the money spent (~ $20). Let me reset and I'll give the dimensions.

<PAUSE for the Repeater to reset >



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Would anyone like any of the dimensions repeated?

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Another antenna that is a good performer for fixed operations is the " American Legion J-Pole ". They are generally available at both the Foothill and Livermore Flea Markets for a reasonable price. Attach it to the top of a 10 foot section of PVC pipe and mount this to a camera tripod. Attach weights to the legs for stability. This makes a very nice fixed station antenna with good gain.

<PAUSE for the Repeater to reset >

BATTERY BASICS

For Emergency Operations, it is highly recommended that you carry two fully charged nicad packs and an extra AA battery case and batteries. The nicads will power your HT for at least the first full day of operations and the AA's will allow you to continue to operate if you can't recharge your nicads. It's also important in cold weather to keep nicads warm, and not exposed on your belt.

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As an alternate or primary power source, use 12-volt, 2 Ahr or higher Sealed Lead-Acid or Gel Cell batteries. They fit in a coat pocket, run an HT all day or power a 2 meter mini-amp for 3 hours at a typical duty cycle. Sealed Lead-Acid or Gel Cell batteries have many advantages. They will allow you to:

- **Operate when other forms of power are not available**

- **Operate longer than with NiCad or Alkaline batteries; 1–3 Ahr batteries are still small enough to be hand carried.**

- **Operate mobile, portable or fixed at a higher output power**

- **Operate at a lower cost; Gel Cells and SLA's are less expensive than NiCads or alkalines.**

- **Operate with fully charged batteries at all times; It's possible to keep these batteries on a "smart" charger continuously.**

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Sealed Lead-Acid batteries are used to power medical diagnostic instruments, alarm systems and Un-interruptible Power Supplies (UPA) just to name a few. Depending on the criticality of the application, some organizations replace their batteries on a regularly scheduled basis well before they are worn out and require disposal as hazardous waste. EC's should write or call local hospitals, explaining how batteries they discard are useful for emergency communications activity. It may be possible to obtain a quantity free for the asking, with no more trouble than signing a hand receipt to satisfy the environmental officer and writing a thank you letter to the hospital administrator. Remember, a hospital's "donation" to your ARES group eliminates their disposal cost.

<PAUSE for the Repeater to reset >

Besides hospitals, and alarm companies, batteries are also available locally at reasonable prices. For example, Halted Specialties usually carries new 12V, 7Ahr batteries for around $20. For those wanting to buy a complete commercial package, HRO sells the "Hot Pocket" which is a 12V, 2Ahr SLA for $78. It comes with its own pouch for attaching to your belt and a wall charger.

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If you decide to use Sealed Lead-Acid or Gel Cell batteries, you'll need a battery charger. 12 V Battery Chargers are available from various sources:

- **Jade Products offers a "Smart Charger" for charging Gel Cells and SLA's. It's $126.95 in kit form.**

- **A & A Engineering also offers a "Smart Charger" for Gel Cells and Lead-Acid batteries. Their charger is $59.95 in kit form or $79.95 finished .**

You can also use so-called "Wall Warts", but the general consensus is that your batteries won't last as long on these inexpensive chargers.

<PAUSE for the Repeater to reset >

Of course, you can always build you own. Two very good articles are the:

June '87 issue of QST entitled "A New Chip for Charging Gelled-Electrolyte Batteries"

This article uses the Unitrode UC3906 "Smart Chip".

March ‘94 issue of QST entitled "A Lead-Acid Battery Charger"

There is also some very good information available on the internet. I have several good URL's that I can pass along if anyone is interested.

<PAUSE for the Repeater to reset >

http://www.unitrode.com	**Unitrode – Home of the UC3906 "Smart Chip"**
http://www.benchmarq.com/	**Benchmarq chargers**
http://www.ibexmfg.com/	**Ibex battery chargers**
http://www.yuasabatteries.com/	**Yuasa "Battery Handbook and Technical Guide"**
http://www.electriciti.com/batteri/pwrson1v.html	**Powersonic batteries**

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AUXILIARY CONNECTIONS

Power cords for connecting to automobile cigarette lighter plugs or gel cell batteries will be needed for extended operation. Be aware that cigarette lighter plug power may be unreliable due to contaminated cigarette lighter sockets. Also, the sockets themselves vary in size, allowing the plug to vibrate loose. As an alternate source of power, however, everyone should still have the capability since they are readily available.

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Be sure to assemble any auxiliary power cords for your HT or small power amps following the standard wiring configuration shown in your respective SPECS or SVECS manual. The configuration that follows is taken from the ARRL ARES Resource Manual. The SVECS manual agrees with this configuration; I’m not certain of the SPECS manual. This configuration uses twin lead 12 to 16 gauge "zipline" with Molex Series 1545, 2-pin polarized connectors and .093 diameter pins. In ARES practice, the

female pins are assembled into the male plug which is attached to the power source, and the male pins into the female receptacle which is attached to the rig. This description will make more sense to you when you have the parts in front of you and are assembling the connector.

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The plug, receptacle and pin sets are available from Radio Shack, Part No. 274-222 and are rated at 12 amps, which is adequate to power small amplifiers up to 50w output. Wiring is simple. The end of the two-pin Molex plug in cross section resembles a little 2-story house with peaked roof. Remember proper polarity by using the word associations "red roof" and "black basement" for the positive and negative terminals respectively. Crimp the wires in place before soldering to ensure a strong connection. After inserting the pins into the plug and receptacle, check the fit of the assembled fitting and reinforce the wires behind the plug and receptacle with heat shrink tubing or tape.

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For re-charging the battery, attach crimp type .187" female tab terminals to fit the male tabs on the battery. The other end of the battery connection will go to a standard cigarette lighter socket. Next, wire a plug onto the leads of a 12-15v, 500 to 900mah charger. A depleted battery can be restored in 4 to 6 hours by plugging it into your car's battery. Gel cells and SLA's larger than 5 Ahr may be left on a 12-14V DC "smart" charger of 1 amp without harm. A word of caution here . . . . . become very familiar with your particular charger and the types of batteries that you intend to charge before leaving a charging battery unattended. A malfunctioning charger can create a very BIG mess very quickly.

<PAUSE for the Repeater to reset >

HAND HELD DUTY CYCLE LIMITS

If you subject compact HTs to frequent full power 5w transmissions of several minutes duration they will overheat and the final power transistors may fail prematurely. Kenwood and Yaesu state in their service manuals that their HTs are rated for 20% duty cycle at maximum RF output, or 30 seconds of transmit to 2 minutes of standby.

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Of the popular 2-meter HT's, Standard does not restrict duty cycle on theirs, rating their amateur hand helds equal in that respect to their aviation, marine, commercial and public safety portables. Unless your HT is a Standard, older Icom or converted

commercial gear, it is best to use your HT mostly on medium or low power for long winded rag chews and restrict full power 5w use to short transmissions to save the finals. If you have a need for high power transmissions of several minutes duration and can't replace or supplement your hand held with a mobile rig, I would advise getting a power amplifier to do the heavy work. Just remember to reduce your HT's output power going into the amplifier, otherwise, you will still burn up your finals.

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ADVICE ON POWER AMPLIFIERS

An excellent way for HT owners to upgrade their portable ARES/RACES equipment is to purchase a power amp. An ideal amp should weigh less than 1 pound, be capable of 10 to 15w output when driven by an HT at 1w, or 20 to 40w output when driven by the same HT at 2 to 3w output. It should draw no more than 8 amps of current at its maximum rated output, enabling it to operate safely from the .093 diameter pin Molex Series 1545 connector or fused cigarette lighter plug.

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An FM-only amplifier without a preamp will be adequate in most cases. A preamp tends to accentuate intermod due to paging transmitters. Unless you equip your portable station with a notch filter and/or cavity bandpass to suppress this intermod, the preamp will serve no useful purpose. Be wary of bargain "no-name" amps you see at hamfests or in discount catalogs. Some are not aligned for the entire U.S. 2-meter band, many lack thermal protection circuitry for over voltage, overdrive or high VWSR or simply have an inadequate heat sink and will overheat and quit.

<PAUSE for the Repeater to reset >

OK, this concludes "Emergency Operations and your HT." Are there any questions?

Web Site Home Page

This page was last updated 04/21/02

The Effects of VSWR on Transmitted Power

By James G. Lee, W6VAT

No matter how long you have been a ham, sooner of later you will be involved in at least one discussion of something called the Voltage Standing Wave Ratio, or VSWR, of an antenna system. There is a lot of good information available on VSWR as well as a lot misconceptions about what it is and what it signifies. Probably the most often misconception is that your VSWR should be as close to 1:1 as possible, otherwise " you won't get out very well." A 1:1 VSWR implies a perfect match between all elements of the antenna system. The only problem is that it is possible to have a low VSWR and still have some very serious things wrong with your antenna system. Other misconceptions such as a high VSWR causing television interference, or other unwanted problems are often heard and can cause unnecessary worry. The concept of VSWR is easy to grasp and its importance in an antenna system does not require an engineering degree to understand.

WHY VSWR EXISTS

Early in electronics you learned that to get maximum power into a load required that the load impedance match the generator impedance. Any difference, or mismatching, of these impedance would not produce maximum power transfer. This is true of antennas and transmitters as well but, except for handie-talkies, most antennas are not connected directly to a transmitter. The antenna is usually located some difference from the transmitter and requires a feedline to transfer power between the two. If the feedline has no loss, and matches BOTH the transmitter output impedance AND the antenna input impedance, then - and only - then will maximum power be delivered to the antenna. In this case the VSWR will be 1:1 and the voltage and current will be constant over the whole length of the feedline. Any deviation from this situation will cause a "standing wave" of voltage and current to exist on the line.

There are a number of ways VSWR or its effects can be described and measured. Different terms such as reflection coefficient, return loss, reflected power, and transmitted power loss are but a few. They are not difficult concepts to understand, since in most instances the are different ways of saying the same thing. The proportion of incident (or forward) power which is reflected back toward the transmitter by a mismatched antenna is called reflected power and is determined by the reflection coefficient at the antenna. The reflection coefficient "p" is simply a measure of this mismatch seen at the antenna by the feedline and is equal to:

$$P = (Z1 - Z_0)/(Z_1 + Z_0)$$

Here $\mathbf{Z_1}$ is the antenna impedance and $\mathbf{Z_0}$ is the feedline impedance. Both $\mathbf{Z_1}$ and $\mathbf{Z_0}$ are complex numbers so "p" is also a complex number.

You remember from elementary AC mathematics that a complex number has a "phase angle" associated with it. The phase of the reflected signal will be advanced or delayed depending upon whether the antenna appears inductive or capacitive to the feedline. If the antenna appears inductive the voltage will be advanced in phase, and if the antenna is capacitive, the voltage will be retarded. The reflective signal travels back to the transmitter and adds to the incident signal at that point.

Thus, any mismatch at the antenna gives rise to a second 'travelling wave' which goes in the opposite direction from the incident wave. When $\mathbf{Z_1 = Z_0}$ the reflection coefficient is zero and there is no reflected signal. IN this case all power is accepted by the antenna and this is the ideal situation where VSWR is concerned. The problem is that this condition is rarely, if ever, achieved and so "p" will have a value different from zero. Note that "p" can have negative values, but in calculating VSWR from the reflection coefficient, only the "absolute value" is used - which is a positive value lying between 0 and 1.

As the two travelling waves pass each other in opposite directions, they set up an interference pattern called a "standing wave". At certain places on the feedline the voltages will add producing a voltage maximum, and at others their relative phase difference will cause a voltage minimum to exist on the feedline. These maximum and minimum points occur 1/4 wavelength apart. In the days when open-wire feedlines were used these points could easily be measured with simple indicators. Coax cable however presents another problem since the "inside" of the cable is not readily available for measurements. Consequently, VSWR measurements on coax are usually made at the transmitter end of the feedline. Therefore you are presented with the VSWR of the entire system which includes all losses associated with the entire system.

INTERPRETING WHAT YOU HAVE READ

Many VSWR meters are calibrated to read FORWARD power as well as REFLECTED power. They may actually be measuring voltage,

and simply have the scales calibrated in power. The important point is to understand what the meter is actually telling you. Assuming for the moment that the VSWR meter contributes no errors, the FORWARD reading is the SUM of the forward power and the reflected power. As a result, it is greater than your actual power output. The REFLECTED power reading is that amount of power which was not initially absorbed by the antenna and has been sent back down the feedline. At the transmitter end it encounters the transmitter output circuitry and is re-reflected back towards the antenna. This happens because you do, in fact, have a VSWR greater than 1:1 as seen by the transmitter. When the re-reflected power encounters the antenna, a portion of it is absorbed and the whole process starts over again.

Ultimately then, most of your signal is eventually absorbed by the antenna. You might be tempted to think that all of this bouncing back and forth would cause "smearing or blurring " of your signal but this is not so. The average transmitted signal appears as a "steady-state" signal to the feedline and antenna. Remember your signal is travelling at a significant fraction of the speed of light. For instance, the velocity of propagation of RG-8/A is 0.66 or 2/3 the speed of light. The speed of light is close to 1000 feet per microsecond, and a dot or voice peak takes milliseconds to complete. If the speed of light were 20 miles-per-hour then the situation would be completely different and we probably wouldn't have radio transmission at all. (Ed. Note, it would be as fast as the mail then.)

Given the reality then that almost all power launched down a feedline reaches and absorbed by the antenna, one has to wonder why VSWR is all that important. The importance is due to the fact that feedlines have losses and, antennas have something called radiation efficiency. They are what make proper interpretation of VSWR important. Power is lost due to feedline attenuation and this loss goes up as the VSWR goes up. The efficiency of an antenna is determined by the ratio of its "radiation resistance" to its "loss resistance". Antenna efficiency can simply described by the following equation:

$$\% \text{ Efficiency} = [R_a/(R_a+R_{loss})] \times 100$$

The radiation resistance is $\mathbf{R_a}$, and $\mathbf{R_{loss}}$ is made up of any associated losses of the antenna such as loading coils and ground systems. How well you "get out" therefore depends more on low losses and efficient antennas than on what your actual VSWR is as the following example will show.

THE EFFECTS OF ATTENUATION ON VSWR

Early in this discussion the statement was made that your VSWR may appear to be very low and yet there could be serious things wrong with your antenna system. Figure 1 shows how this can happen. The curves in the figure represent the forward and the reflected voltage on an antenna which has a feedline loss of 3 dB. and a reflection coefficient of p=0.5. In this example the actual value of voltage is inconsequential and can be considered to be "E". We are only interested in relative values of "E" in any case. The length of the feedline is also arbitrary since we are only concerned with its total loss between transmitter and antenna.

Figure 1

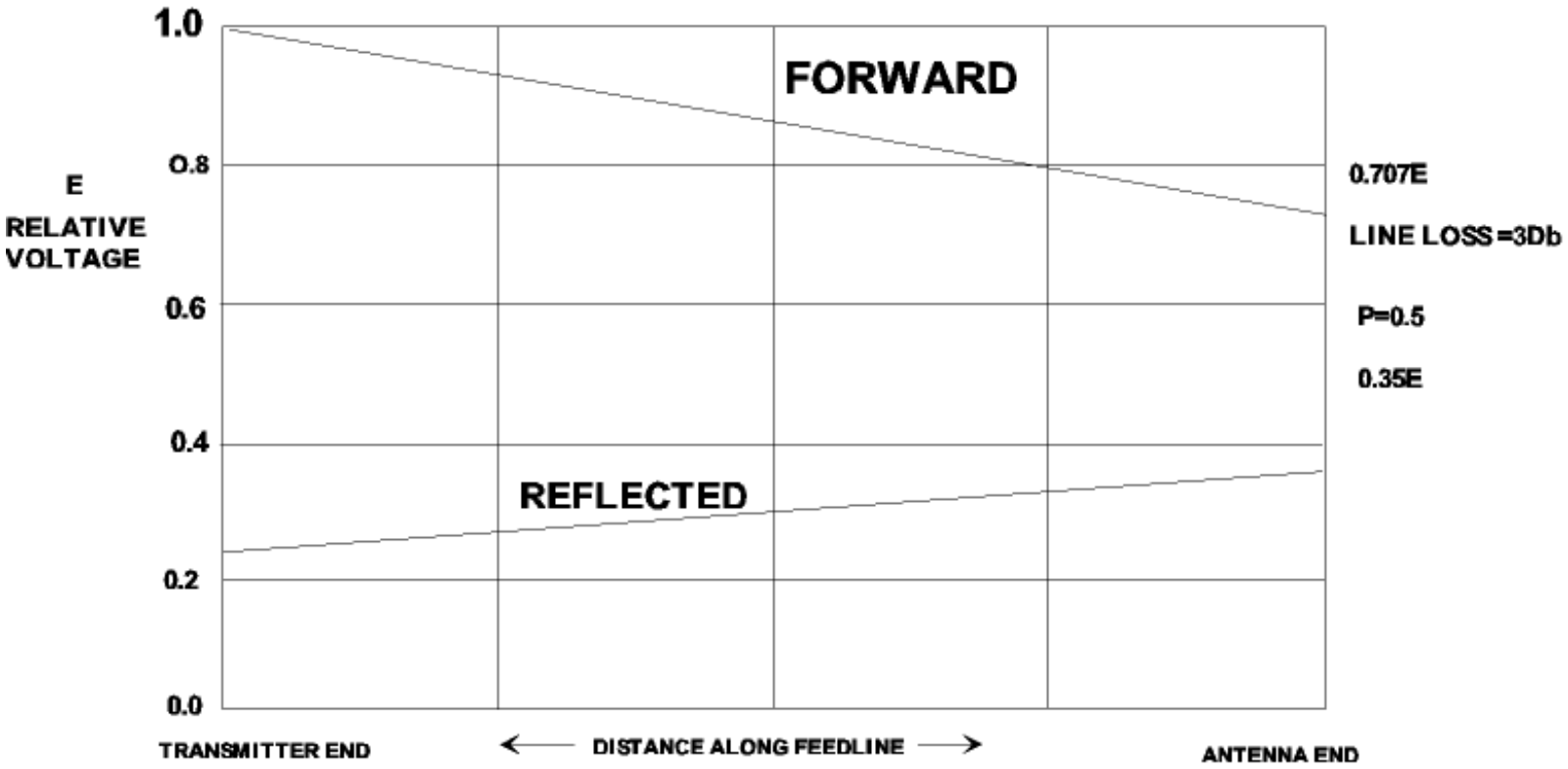

FIGURE 1: EFFECT OF LINE LOSS ON SIGNAL AMPLITUDE

The signal voltage "E" starts out at full value -1.0 E - on the feed line and is attenuated at a 3-dB rate. This means that the voltage will only be 71% - or 0.707E - when it reaches the antenna terminals. Remember that while 3-dB is a factor of two for power considerations, power is proportional to E-squared, consequently E will be only 0.71e when it arrives at the antenna input. The top curve in Figure 1 shows the FORWARD voltage decay as it travels down the feedline to the antenna input.

Since the antenna in this example has a reflection coefficient of 0.5, this means that 1/2 of the incident voltage will be reflected back down the feedline. This value is (0.5 X o.71E) or 0.35E volts. The feedline has no way of knowing which way signals are traveling, so this reflected voltage will suffer the same 3-dB attenuation on the return trip. When it arrives back at the transmitter end of the feedline its value is only (0.71 X 0.35E) or 0.25 volts. The VSWR meter sees this value and since

$$\mathbf{VSWR=(E_{fwd} + E_{ref})/(E_{fwd} - E_{ref})}$$

the VSWR meter will read 1.67:1

That value of VSWR is guaranteed is to make almost everyone happy, but your antenna system is not very good. The 3-dB loss down the feedline means only 1/2 of your output power reaches the antenna, and if your antenna has significant losses, something less than 1/2 of your power will be radiated depending upon how bad the losses really are. If for instance, the loss resistance equals your radiation resistance, the antenna is only 50% efficient meaning only 1/4 of your output power is actually radiated. Yet that reading of 1.67:1 looks fine. A reflection coefficient of p =0.5 means your antenna is not well matched to the feedline. VSWR can be calculated from the reflection coefficient by the following:

$$\mathbf{VSWR = (1+p)/(1-p)}$$

Using this formula shows your VSWR at the antenna is 3:1, quite a different value than your VSWR meter reads. The difference in the input and output VSWR values is due to the loss introduced by the feedline. Figure 2 shows how this loss can cause you to get a different VSWR depending upon where you measure VSWR in a feedline. You can measure VSWR at the antenna end of the feedline, but it is usually impractical to do.

Figure 2

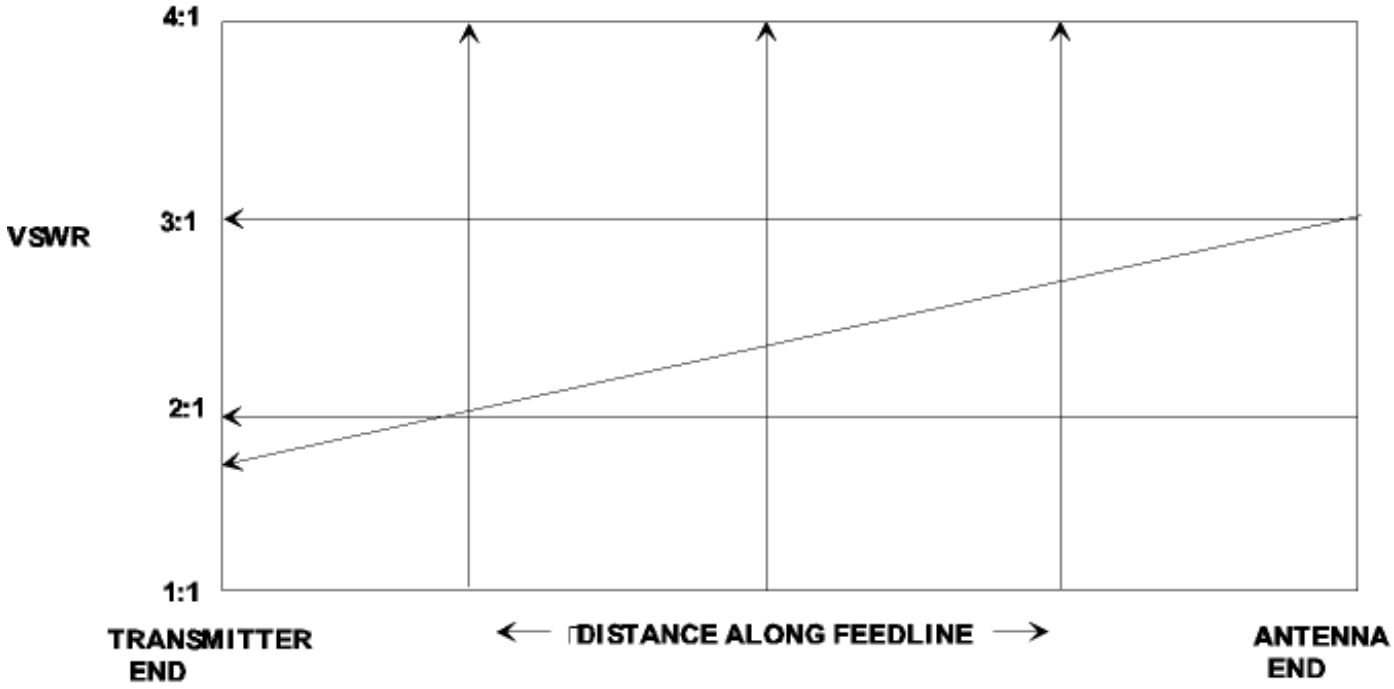

FIGURE 2 EFFECT OF LINE LOSS ON VSWR

You can use 1/2 wavelengths of coax between your VSWR meter and the antenna because a 1/2 wavelength of cable repeats the impedance it sees. The only problem is that you are introducing other possible elements into your measurements. But let's assume that your VSWR measurement at the feedline is reasonably close to what is actually occurring on the feed line, and that your feedline losses are not great. The burning question still is "how good or bad is the VSWR reading?"

VSWR AND TRANSMITTED POWER

Let's assume you have an efficient antenna, fed with a low-loss feedline so that the VSWR meter at the transmitter gives you a true reading of 1.65:1. There is no real reason to try to lower it, in fact the same would hold true if the reading were 2:1. Figure 3 is a chart showing the equivalence of VSWR to RETURN LOSS(dB), REFLECTED POWER(%) and TRANSMISSION LOSS(dB). Return loss is related to reflection coefficient by the equation:

$$\textbf{Return Loss} = -20\log_{10}(p)$$

It is just another way of measuring VSWR. For example, with a perfect 1:1 VSWR there would be no reflected power consequently the return loss on the feedline would appear to be infinite. A short or open circuit at the antenna is the worst case scenario since the reflection coefficient would be p =1.0. All incident power would be reflected, and with a lossless feedline the return loss would be 0-dB. this is what the RETURN LOSS (dB) column refers to

The most informative columns in Figure 3 are the REFLECTED POWER(%) and the TRANSMISSION LOSS(dB) columns since they provide an answer to our question of whether further reduction of VSWR is worthwhile. Figure 3 shows that for a VSWR of 1.65:1 the reflected power is only 6.2% of the incident power, and the transmission loss is only 0.27 dB. In more familiar terms, if you count an S-unit as 6 dB, then the 0.27 dB loss is only 1/22 of an S-unit. A reduction of the VSWR to 1.5:1 would provide only a 0.09 dB reduction in transmission loss. This is not worth the effort it would take to achieve such a miniscule increase in power.

Figure 3

VSWR	Return Loss (dB)	Reflected Power (%)	Transmiss. Loss (dB)	VSWR	Return Loss (dB)	Reflected Power (%)	Transmiss. Loss (dB)
1.00	oo	0.000	0.000	1.38	15.9	2.55	0.112
1.01	46.1	0.005	0.0002	1.39	15.7	2.67	0.118
1.02	40.1	0.010	0.0005	1.40	15.55	2.78	0.122
1.03	36.6	0.022	0.0011	1.41	15.38	2.90	0.126
1.04	34.1	0.040	0.0018	1.42	15.2	3.03	0.132

1.05	32.3	0.060	0.0028	1.43	15.03	3.14	0.137
1.06	30.7	0.082	0.0039	1.44	14.88	3.28	0.142
1.07	29.4	0.116	0.0051	1.45	14.7	3.38	0.147
1.08	28.3	0.144	0.0066	1.46	14.6	3.50	0.152
1.09	27.3	0.184	0.0083	1.47	14.45	3.62	0.157
1.10	26.4	0.228	0.0100	1.48	14.3	3.74	0.164
1.11	25.6	0.276	0.0118	1.49	14.16	3.87	0.172
1.12	24.9	0.324	0.0139	1.50	14.0	4.00	0.18
1.13	24.3	0.375	0.0160	1.55	13.3	4.8	0.21
1.14	23.7	0.426	0.0185	1.60	12.6	5.5	0.24
1.15	23.1	0.488	0.0205	1.65	12.2	6.2	0.27
1.16	22.6	0.550	0.0235	1.70	11.7	6.8	0.31
1.17	22.1	0.615	0.0260	1.75	11.3	7.4	0.34
1.18	21.6	0.682	0.0285	1.80	10.9	8.2	0.37
1.19	21.2	0.750	0.0318	1.85	10.5	8.9	0.40
1.20	20.8	0.816	0.0353	1.90	10.2	9.6	0.44
1.21	20.4	0.90	0.0391	1.95	09.8	10.2	0.47
1.22	20.1	0.98	0.0426	2.00	09.5	11.0	0.50
1.23	19.7	1.08	0.0455	2.10	09.0	12.4	0.57
1.24	19.4	1.15	0.049	2.20	08.6	13.8	0.65
1.25	19.1	1.23	0.053	2.30	08.2	15.3	0.73
1.26	18.8	1.34	0.056	2.40	07.7	16.6	0.80
1.27	18.5	1.43	0.060	2.50	07.3	18.0	0.88
1.28	18.2	1.52	0.064	2.60	07.0	19.5	0.95
1.29	17.9	1.62	0.068	2.70	06.7	20.8	1.03
1.30	17.68	1.71	0.073	2.80	06.5	22.3	1.10
1.31	17.4	1.81	0.078	2.90	06.2	23.7	1.17
1.32	17.2	1.91	0.083	3.00	06.0	24.9	1.25
1.33	17.0	2.02	0.087	3.50	05.1	31.0	1.61
1.34	16.8	2.13	0.092	4.00	04.4	36.0	1.93
1.35	16.53	2.23	0.096	4.50	03.9	40.6	2.27
1.36	16.3	2.33	0.101	5.00	03.5	44.4	2.56
1.37	16.1	2.44	0.106	6.00	02.9	50.8	3.08

Further examination of the chart shows that a VSWR of 2.6:1 results in only about 1 dB of transmission loss. A high VSWR of 6:1 shows just a 3 dB transmission loss, but this is 1/2 an S-unit. You will still be getting out but this is becoming a significant loss. Your feedline will be dissipating more power than it should, and there may be other serious things wrong with your antenna system.

Throughout this article you've noticed the use of the term "antenna system". The word "system" means you must pay attention to other things besides just the VSWR and your power output. Each component of your antenna system must be optimized to get the best results. Many factors must be considered such as operating frequencies, bandwidth requirements of the antenna system, heights, and directivity, all of which affect its efficiency. Since the height of your antenna, and your operating frequency determine both the length of the feedline and its losses the interfaces become very important. So there are a number of trade-offs which must be considered when you contemplate putting up a good antenna system, but these are tales for other times.

You can build or buy your own VSWR meter, but make sure that you understand what it is measuring and what it is really telling you. Then once you are satisfied that you have put up a really efficient antenna, fed with a low loss feedline, you can sleep well knowing that to try to reach the ultimate 1:1 VSWR is only an ego trip. As a rule of thumb, any accurate VSWR reading under 2:1 is probably not worth the effort to achieve if the other elements of your antenna system are the best you can make them. In fact you might be surprised to find that you really do have a low VSWR when you put up the best antenna and feedline you can. There is an old saying in ham radio that "a dime in the antenna is worth a dollar in the transmitter any day". Try it and see if you don't agree. **-30-**

Editors note:

W6VAT has a lot of good points, and careful attention should be paid to what is covered in his article, as it can make a difference in your signal. A case in point involves a ham club that I belonged to many years ago. They had just gotten the license for their repeater, and the only antenna that was available was a commercial antenna fed with ancient heliax. The antenna had only a small amount of reflected power and it seemed to get out well. Everyone was happy, and all was well with the world.

Until my boss who was a ham and I took the liberty of checking out the repeater with the equipment from the two-way shop. When I disconnected the RG-8A pigtail, water poured from the heliax for about 5 minutes. But the Bird showed no reflected power with a 10 watt element and 50 watts forward. This was met with extreme disbelief when the club members were told of the water cooled coax. No one wanted to spend the money to replace the antenna and coax. However, since we had a very heavy rain just before the meeting, duplication of the waterfall was easily achieved.

To the extreme displeasure of the older members who wanted to "patch the feedline and connections" a motion to spend the money to replace the old antenna and coax with new everything. After the purchase of a Ringo and some 1" heliax and the installation of same, the repeater range was tripled. The grumblers did not like the Ringo, as "it was not good enough". And, after I had left the area, they spent nearly two hundred dollars for a commercial antenna cut for the frequency, which they purchased from a club member who ran a two way shop and was adamant that the Ringo was no good. They were totally destroyed to find out that the high dollar antenna gave them no more range than the Ringo.

The moral to this story is things are not always as they seem, especially when it comes to VSWR, coax, and antennas that seem to work on frequencies that they shouldn't. Never take anything for granted, especially when it is your RF going up the flue. -30- *K5CNF*.

Send mail to webmaster@antennex.com with questions or comments.
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Last modified: July 04, 2004

ANTENNA TOPICS

Antenna Software and more

GW0TQM's Mag Loop page

PA3HBB's site - Magnetic loops and more!

End feds, doublets etc - nice reading

End fed antenna ideas

Nice antenna designs for HF and VHF (PA3HBB)

A Short dipole for 80M

CT1ETT's keys, mag loop and rigs

Excellent site for antennas!

A Simple LED SWR Bridge

The Cobwebb HF Antenna

The Random Wire Vertical

Notes on the Random Wire Vertical

A Light Duty Mast

A Mini ATU

The OHR WM2 Wattmeter

The Broken Quad

Groundplanes by G3PTO

SM0VPO's Mini HF Vertical

N3GO's J-pole

A Multiband loop

How to make mobile/portable whips

A Resistive SWR Bridge for QRP Use

The Cross Field Antenna

The Lattin 5 Band Antenna

A practical 160m antenna

The Square Pole

Try an Inverted-L

The G5RV multiband HF Antenna

More on The G5RV multiband HF Antenna (incl. 160m)

The W3EDP HF antenna

The 'Shorty' mini-dipole for 14mHz

Another Antenna Idea (A 3-band dipole)

A Shortened 7mHz Dipole

Modified Z Match ATU

A Single Coil Z-Match ATU

An Antenna idea for the Smaller Garden

A Magnetic Loop

The Rockloop Compact HF Antenna

An Antenna for Portable Use

The Autek Antenna Analyzer

The MFJ259 Antenna Analyzer

Click To View SM0VPO's Pages For More Antenna Ideas

W4RNL's WEB SITE - MINE OF INFORMATION ON ANTENNAS!

Click for the web site

Another great site for antennas!

Click here

Frank, G3YCC

Back to the first page

ARRL The national association for AMATEUR RADIO

The CarChip™ ODB II Scan Tool is @ www.provantage.com

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Text-only/Printable

Antenna Projects

ARRL Technical Information Service page · TIS Menu page

Introduction

QST, through the years, has had more articles on antennas than any other subject. This is probably because your antenna, although it can be the least expensive, easiest to construct component of your station, can have the greatest effect on performance.

The main feature here is a menu of antenna pages divided into type and use. There are all kinds of antenna projects you can build for home, portable and mobile use, HF and VHF. Go to the Technical Information Antenna Projects Menu to find what you need. For the new-comer there is a special Beginners Section covering both HF and VHF antennas.

Additional resources

The ARRL Antenna Book
ARRL's Wire Antenna Classics
More Wire Antenna Classics Volume 2
Vertical Antenna Classics
Antenna Compendium Vol's 1 through 6
QST Magazine

The ARRL has an extensive catalog of books and materials related to Amateur Radio.

ARRL Products: Interference/DF
(More)

AC Power Interference Handbook -- New insights into the causes, effects, locating and correction of power-line and electrical interference. 2nd Edition.

The ARRL RFI Book -- Practical Cures for Radio Frequency Interference.

Transmitter Hunting -- Radio Direction Finding Simplified

The RSGB Guide to EMC -- Tackle RF interference problems and understand the underlying causes.

Technical Information Antenna Projects Menu

HF Antennas VHF/UHF Antennas

Beam/Yagi
Dipole
J-Pole
Loop
Mobile
Quad
Sloper
Trapped Antennas
Vertical
Other HF Antennas

Beam/Yagi
J-Pole
Mobile
Vertical
Quad
Other VHF Antennas

Information for HF and VHF beginners

General Antenna Info Articles:

- HF Amplifiers versus Antennas--One Ham's Opinion (39,999 bytes, PDF file)
  QST November 1998, pp. 54-56
  The solution to "getting out better" is not necessarily "more power".
- Low Power, Crummy Antenna (293,108 bytes, PDF file)
  QST December 1998, pp. 57-59
  A major part of the fun of amateur radio is putting together a station that works from stuff you have just laying around. Antennas lend themselves to this practice nicely.
- Antenna and Tower Safety (15,694 bytes, PDF file)
  QST June 2001, p. 91
- "What's Up Top"
  QST June 1960, pp. 38-40
  Crud and rust can be a real detriment to the performance of your antenna. Here are some procedures to make it look (and work) like new again.
- Stacking Yagi Antennas
  By what distance should stacked Yagi's be separated?
- Beam Talk for the Layman (263,249 bytes, PDF file) MEMBERS ONLY
  QST July 1958, pp.35-37
  Step-by-step tuning of a gamma matched Yagi antenna and other thoughts on installation.
- Sound-Card Antenna Measurements and Other Useful Techniques (1,560,065 bytes, PDF file) MEMBERS ONLY
  QEX Jan/Feb 2002, pp. 33-46
  Measure antenna and receiver performance and record results, even when you are not present, using your computer's sound card.
- Tower and Antenna Wind Loading as a Function of Height (192,267 bytes, PDF file)
  QEX July/August 2001, pp. 23-33
- A Low-Loss VHF/UHF Bias Tee (279,405 bytes, PDF file) MEMBERS ONLY
  QEX May/June 2002, pp. 52-54
  A simple circuit that lets you superimpose DC control voltage onto the transmission line without altering the RF characteristics.
- Tower and Antenna Wind Loading as a Function of Height (192,267 bytes, PDF file)

QEX July/August 2001, pp. 23-33

- An Inexpensive External GPS Antenna (253,745 bytes, PDF file)
  QST October 2002, pp. 36-39
  An easy to build antenna for your GPS unit for better reception.

Web Links:

- Stacking, Phasing and Matching Yagis
  This is a synopsis of a talk presented to the Sydney VHF DX Group on Tuesday March 16th 1999 by Gordon McDonald VK2ZAB
- How to Become An Antenna Guru
  What Is An Antenna, Antenna Impedance, Transmission Lines, Radiation Pattern, Polarization, and more.
- The Antenna Elmer
  Here you will find information about wire antennas as well as directional beams.
- KB0YKI's Radio Zone
  On-line antenna designers, rooftop tower plans, gamma matches, and more
- Amateur Radio
  My personal interest in amateur radio focuses on research into and education about antennas and antenna modeling.
- EZNEC Antenna Software by W7EL
  EZNEC 3.0 and EZNEC pro 3.0 are full-featured 32-bit Windows applications
- NEC4WIN95VM v3
  Antenna Simulation Software for Windows
- Polar Plot by Bob Freeth (G4HFQ)
  A program that lets you see what the polar diagram of your relatable beam antenna actually looks like where it is operating.
- AC6V's Homebrew Antenna Links
- Antennas, L. B. Cebik, W4RNL
  Theory (light) and practical information on all kinds of antennas.
- Bill, NJ7P Coaxial Cable Searchable Database
- Lines of Magnetic Declination for North America 1990
  Useful to find true north for aligning your antenna
- FM Broadcast Directional Antenna for 87-93MHz
  Build this ideal antenna for receiving distant Public Broadcasting stations
- KF4VDF's Antennas
  Diagrams, measurements and calculators for designing and building all sorts of antennas
- G4FGQ's Software
  New, original, high-quality, radio engineering/modeling programs. Not freeware -- just free gifts -- please help yourself -- no catch.

Article Reprints:

Note: *Reprints are available from the Technical Department Secretary, ARRL, 225 Main Street, Newington, CT 06111. Article copies must be prepaid $3 for ARRL Members, $5 for non-members. You may order by telephone (860-594-0200) and pay by credit card. Reprints can only be sent by mail and cannot be sent by or attached to e-mail.*

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HAM RADIO SPECTRUM
REFERENCE AND DIRECTORY

Host Your Ham Site With Us | You can have (email) anything@yourcall.com (site) www.yourcall.com

ANTENNAS

Below are links to some Antenna pages and sites. If you have an antenna page please send your link to us so we can share it with others. If you are looking for some info you can't find, let me know and I'll try and help you locate it. Be sure to check back as we are updateing daily. If you have A link or info, send it for listing.

40 Meter 2 Element Parasitic Delta Loop

A simple and inexpensive way to achieve substantial gain and reasonable F/B ratio.

Amateur radio antennas

Super Linear Loaded Inverted V ,Copper Cactus (Multi-Band J-Pole),The Stacked-J (Double High J-Pole), and more

---

Antenna programs

programs for all the common antennas, hf, vhf ,uhf and some experimental designs

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2 meter j-pole

A simple J-pole antenna for 2 meters

FEATURED SITES
- Java Radio
- Dx Spotting
- Space Weather.com

GENERAL INFO:
- Ham Radio History
- Famous Hams
- HF Band Plans

SPECTRUM CATAGORIES
- Antennas
- Clubs
- Callsigns
- Digital
- DX
- Exams
- Hamfest
- Homebrew
- Humor
- Manufacturers
- Mods
- Personal Pages
- Publications
- Repeaters
- Scanners
- Software

40 meter half square $29.95

eHobbies has remote control plains, cars, rockets, helicopters and much more.

All about baluns

Theory and construction

Srewdriver moble antenna mount

A very nice mount from K9ROY

DX dual dipole

A wire yagi

VHF/UHF Long Yagi Workshop

Designs for VHF/UHF yagis

Quickie Vertical

---

7 ELEMENT LOGCELL YAGI

Claimed to be better than a regular yagi but on a shorter boom

---

Fractal quad for 10 meters

It's small, high performance, and unencumbered by electronic parts, transformers, discrete loads, and so on. And a 50 ohm feed

---

Rotatable Dipoles

6 meters, 20 meters and multiband

NB6ZEP antenna

A modified 20 meter Extended Double Zep, 40 thru 10 meters (with tuner)

Cubicle quad antenna calculator

This page makes calculates all dimentions for quad antennas, just enter the frequency. With detailed plans.

1270Mhz Yagi Antenna

Build your own antenna for the 23cm band (1250Mhz - 1280Mc) using some aluminium and this simple design

---

2 EL. Quad for 10-12-15 meters

Simple disign useing bamboo poles, of course fiberglass, pvc, etc could be used.

---

80 METER FRAME ANTENNA

A nice antenna for limited space

ANTENNA ROUND UP

ANTENNA'S FOR 6, 10 AND 2 METERS

K1TTT Technical Reference

Links to antenna resourses

Grounding is key to good reception

Discussion on grounding.

An Attic Coaxial-Cable Trap Dipole for 10, 15, 20, 30, 40, and 80 Meters

A good antenna for HF operation on 6 amateur bands at a QTH with restrictive covenants that prohibit outside antennas.

Clif's Ham Radio Page

Summary of experience with restricted space antennas.

No tuner required

W6RCA's No-Tuner, All-HF-Band, Horizontal, Center-Fed Antenna

---

Folded Dipoles

Dimintions for the folded dipole antenna

160/80 meter recieving loops

Construct this loop for a lower noise receiving antenna

The Gallery of the EME Arrays

Huge picture gallery

---

2 METER HALO ANTENNA

A small horizontal loop

Six Meter Antennas

A 3 ELEMENT YAGI and A 6 meter halo

Antanna calculators

Some great antenna stuff here

Antennex

An antenna magazine

Coax cable table

Loss,velocity facter, etc

The Gain of an Endfire Array

This article looks at the maximum theoretically possible directive gain that could be obtained with an endfire array of given length or number of elements.

Ground-mounted Verticals

Designed verticals for you to build

KB1GW's collection of Beverage Antenna Information

Rhombic Antenna Home Page

Expanations and photos.

A Simple Seven Element Yagi Antenna

Quickly calculate the dimensions for a seven element Yagi-Uda Antenna. Simply enter the freqency in Megahertz and the script will do the rest.

THE ST. LOUIS VERTICAL

The St. Louis Vertical (SLV) offers portable enthusiasts an easy-to-build, easy-to-use antenna.

Standard G5RV Information

A discussion of the G5RV

The Super J-Pole Antenna(Collinear Design)

Javascript for calculating the lengths of tubing to be used for the construction of a Super J-Pole.

Transmission Lines and Matching Systems

Explanation

Rhombic Antenna Home Page

Very nice explanation with pictures

Amateur Radio RF Safety Calculator

Enter your gain, distance, etc and this page will let you know if you comply with F>C>C> regulations

---

Quibcal Quad

Plans and pictures to build your own

A Small Yagi for 50 MHz

Understanding electromagnetic fields and antenna radiation takes (almost) no math

---

Build a portable 10 meter yagi

Phased beverage arrays on 160M, 80M and 40M

Coaxial Traps

RF calculators

EME gallery

pictures of eme antennas

---

haarp

HIGH FREQUENCY ACTIVE AURORAL RESEARCH PROGRAM H.A.A.R.P.

RF GROUND SYSTEMS

Everything you always wanted to know about rf grounding

---

national lightning safty institute

Stacking Yagi Antennas

Using and Understanding Decibels

---

2 meter yagi

---

wire antennas

---

antenna

W6RCA's No-Tuner, All-HF-Band, Horizontal, Center-Fed Antenna

---

Projects and Info Center

Antennas & Transmission Lines

Smart Antenna Systems Tutorial

A smart antenna system combines multiple antenna elements with a signal-processing capability to optimize its radiation and/or reception pattern automatically in response to the signal environment.

THE $4 SPECIAL

Sure, you can find "all-band wire antennas" for sale in the back pages of Ham magazines costing $100 or more. But beware: Marconi spins in his grave everytime a ham buys an aerial instead of building it!

The Stacked J-Pole

The Optimized Wide Band Antenna

Notes regarding stacked installations of antennas

Coaxial Collinear hints from K7ITM

How to Become An Antenna Guru

Using these simple relationships, you can do antenna calculations with a wave of the hand and sound like a pro

---

Bazookas and more

Crossed-Field Antenna References

The antenna elmer

How to build a 432 Mhz Quagi

AM Antennas

If you are not satisfied with your AM reception, you are probably wondering "is there anything I can do to improve it?" Well, you have come to the right place!

ADDITIONAL RESOURCES

- hf Frequency allocations for swl
- Wordwide grid squares/zones
- HF Nets
- Country List
- ARRL, The premier site for hams worldwide
- The F.C.C.

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Web Hosting: $12.95/month

KU4AY Web Hosting offers full featured Web Hosting accounts starting at $12.95 with no setup fees and unparalleled OC3 connectivity. Imagine your own ham radio web site with "www.yourowncall.com as your web site and anythingyouwant@yourowncall.com as your email address. KU4AY Web Hosting Solutions

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1434 High Mesa Drive
Knoxville, TN 37938-4443
cebik@cebik.com

Amateur Radio

Amateur Radio is a communications service consisting of operators licensed and regulated in the United States by the Federal Communications Commission. Amateur radio operators, by regulation engage in emergency and other public service communications, maintain technical skills, and foster international good will via communications.

My personal interest in amateur radio focuses on research into and education about antennas and antenna modeling. The listings on this page link to some past and present research (and to some past research updated before posting on these pages). Since these are essentially working notebooks and not polished articles (for the most part), they may contain typos, misspellings, and a few grammatical infelicities. Moreover, they are subject to revision and updating whenever I discover something more accurate, more useful, or more interesting.

Tales and Technicals:
A Little History and a Lot of Antennas

From time-to-time, I shall post some yarns, mostly taken from my collection of old books, manuals, magazines, and handbooks. I shall also occasionally post from the pages of my notebooks some technical information that may be of use to fellow and prospective hams. To keep this index from being too long to use, I have placed many items in collected groups. So be sure to check the listings that appear collect items together from time to time for additions. Since these are notes and some reprints of casual articles, there will be considerable overlap in places--and many large gaps in other areas. Nevertheless, let's begin with these items:

A Little History, a Little Humor, and a Little Seriousness

- **The Wouff-Hong and The Rettysnitch: Lost Traditions?**(02-16-1998)

- The Wouff-Hong, the Rettysnitch, and the What? The Uggerumph (04-27-2000)
- 10-10 and Lindbergh (02-10-1998)
- Q-signals in 1930 (02-10-1998, 01-08-2001)
- The RST Standard of Reporting (10-02-1997, 12-10-2001, 04-19-2004)
- A Time to Reminisce (02-21-1998)
- To the Sea Again (08-21-2000)
- Why Call "CQ"? (06-20-1998)
- Everything Old Is New Again (10-28-1998)
- A Ten-Minute Timer That Just Won't Quit (01-29-2000)
- Understanding and Producing Scientific and Technical Articles, Proposals, and Reports (07-28-1999)
- On Socio-Cybernetic Modeling (08-20-1999)
- PVC Gurneys for All Occasions (12-09-2002)
- Thoughts on Caring for a Terminally Ill Loved One (11-04-2002)
- In Memory of Jean (12-08-2002)

Antenna Modeling Software Notes

- Antenna Modeling Programs (01-23-2003)
- Notes on Modeling Antenna Elements (05-10-1998
- Notes on Modeling and Convergence Testing (06-23-1997)
- Modeling Loads and Transmission Lines Parts 1-4 (12-29-1997)
- Modeling Trap Antennas (04-02-1997)
- Preliminary Notes on Trap Placement (07-11-1997)
- Systematic Trap Modeling (07-12-1997)
- Why I Use dBi Mostly (06-23-1997)
- Some Pitfalls of Careless dBd-ing (06-03-1999)
- Getting the Most Out of Antenna Patterns (07-17-1997)
- Reading a NEC Deck (03-30-1997)
- NEC-4 vs. NEC-2 With Stepped-Diameter Correction and Autosegmentation (04-06-1997)
- Notes on Two Limitations of NEC-4 (04-10-1997)
- Multi-Diameter Dipoles: MININEC vs. NEC-4 (09-19-1997)
- Close-Spaced Wires: MININEC vs. NEC-4 (09-24-1997)
- Notes on Boom Effects with Short 3-Element 146-MHz Yagis (10-22-2003)
- Notes on Fat-Wire Dipole Convergence (03-11-1998)
- 1 MHz Ground-Wave Analysis: A Comparison Among MININEC and NEC Modeling Implementations (02-23-2001)

- **The Tee-Match: NEC Illusions and MININEC Realities** (01-07-1998)
- **Three-Element Yagi Models: Standards of Comparison** (06-21-1997)
- **NEC-2 and NEC-4: Reading Trends with Caution** (09-12-1998)
- **Modeling and Understanding Small Beams** (07-25-1999)

Practical Antenna Notes: Lower HF (Mainly) Vertical Antennas

- **Self-Contained Vertically Polarized Wire Antennas: A Family Album** Parts 1-6 (05-15-1999, 01-14-2003)
- **Voltage Feeding SCV Loops** (08-14-1998)
- **The 40-Meter Bobtail Curtain as An All-Band Wire Antenna** (05-27-2003)
- **Notes on All-Band Use of Vertical-Plane Deltas** (10-13-1997)
- **Verticals Without Vertigo** (05-13-1999)
- **Some Techniques for Building a QRP Field Vertical** (05-24-2002)
- **The 5/8-Wavelength Mystique** (04-17-2001)
- **Some Facts of Life About Modeling 160-Meter Vertical Arrays** Parts 1-5 (08-01-2001)
- **Models with Buried Radials: A Small Compendium** (08-31-2004)
- **Pursuing the (Nearly) Perfect Parasitic Vertical Array for 160 Meters** Parts 1-2 (01-01-2002)
- **Half-Length 80-Meter Vertical Monopoles: the Best Method of Loading** Parts 1-5 (05-12-1997)
- **Vertical Dipoles and Ground Planes: What Antenna Modeling Reports** (09-22-1998)
- **A Modeling Perspective on "Ground" Planes** Parts 1-5 (06-06-1997)
- **The Insulated Radial Question** (05-01-2003)
- **Where Do I Hang My Hat?** (10-31-1998)
- **The 40-Meter 3-Way Special** (07-20-1997)
- **3 Wires = The Whole World** (02-05-1999)
- **A Triangle for the Short Vertical Operator** (02-05-1999)
- **An 80-Meter LPMA: A Design Idea and a Modeling Dilemma** Parts 1-2 (05-01-2002) (PDF)

Practical Antenna Notes: Lower HF (Mainly) Horizontal Antennas

- **Just Tall. That's All. Some Principles of Portable Antennas to Strive For** (08-29-2003)
- **Wi're We Using Wire?** (01-19-1998)
- **A Note on Wire Size and Material** (08-08-1998)
- **The Zig-Zag Dipole-Doublet** (01-01-2000)

- **Unfolding the Story of the Folded Dipole** (03-12-2000, updated 12-04-2001)
- **Modeling the T2FD** (06-07-1999)
- **Notes on the Terminated Wide-Band "Folded Dipole"** (06-01-2000, 07-19-2000, 02-13-2004)
- **The G5RV Antenna System Re-Visited** Parts 1-3 (05-21-2003)
- **The "Ideal" Back-Up Antenna for 80-20 Meters** (04-14-2000)
- **Some ABCs of A-B-C: Notes on Triangles of Doublets** (02-15-2003)
- **Resurrecting the Y-Doublet** (05-16-2003)
- **Notes on All-Band Use of Horizontal-Plane Loops** (10-28-1997)
- **A 40-Meter Star** (04-13-2003)
- **Horizontally Oriented, Horizontally Polarized Large Wire Loop Antennas** (03-29-1999)
- **Vertically Oriented Multi-Band Antennas** (05-20-2000)
- **The All-Band Center-Fed Inverted-L** (03-06-1999, 06-14-1999)
- **To Trap or Not to Trap** (12-20-1997)
- **Some Notes on NVIS Cloud Burners** (11-05-2002)
- **NVIS: From the Backyard to Professional Installations** (05-08-2003; 05-24-2003)
- **Vertically Radiating Horizontal Antennas** (03-15-1997)
- **The IL-ZX Antenna for 40 Meters** (06-30-1997, 12-27-2001, 09-27-2003)
- **Notes on Full and Shrunken 40-Meter Quads** (12-30-1997)
- **Some Notes on Lower-HF Wire Beams** (09-07-1997)
- **Notes on a Wire Beam for 80 and 75 Meters** (04-09-1997)
- **A Note on Substituting Wire Elements in Lower HF Arrays** (03-10-2000)
- **80-Meter Wire LPDAs** (04-01-2002)
- **Some Notes on Antenna Bandwidth** (01-22-1997)
- **40 + 30 = 50 (Not 70)** (10-23-2001)
- **Short Beams and Operating Bandwidth** (10-05-1997)
- **Don't Be Phased By Phasing** (02-04-1998)
- **My Top 5 Backyard Multi-Band Antennas** (08-11-2004)(PDF)

Practical Antenna Notes: Upper HF (Mainly): Yagis and Relatives

- **Horizontal and Vertical Yagi Orientation** (10-13-1998)
- **Lobe Formation With Height Increases in Horizontal Antennas** (04-30-1999)
- **Modeling 6 Long-Boom Yagis** (07-20-1998)
- **What Can We Expect From a 2-Element Beam?** Parts 1-5 (05-09-1997)
- **Mastering Some Mysteries of 2-Element Beams** Part 1 (05-19-2001)
- **Mastering Some Mysteries of 2-Element Beams** Part 2 (05-24-2002)

- **The Monoband Log-Cell Yagi Revisited** Parts 1 to 4 (07-14-2000)
- **The G4ZU Bird Yagi** (04-02-1997)
- **Coils, Linear Loads, and Capacity Hats: An Overview of Small Loaded Yagis** (01-17-1998)
- **Director/Driven Element 2-Element Yagis: Some Ideas for 12 and 17 Meters** (07-25-1999)
- **Using Moxon Rectangles for WARC-Band Antennas**
  ***Part 1*: Some 17-12 Meter Ideas** (05-01-2000, 08-04-2000)
  ***Part 2*: Some 30-17-12 Meter Ideas** (06-01-2000)
- **A Truly Portable Moxon Rectangle for Nearly No-Tool Field Assembly** (02-01-2003)
- **Designing Moxon Rectangles by Equation and by Model** (10-01-2000)
- **Moxon Rectangles (with Dimension Calculator)**: 16 Items (12-04-2003)
- **The Elusive Moxon Nest** (06-01-2003)
- **Moxon-Modifying the C3-Type Tri-bander** (06-01-2002)
- **A Tale of 4 Beams: The X, the Hex, the Square, and the Rect** (09-01-2000)
- **Understanding the N0KHQ Coax Square** (12-08-2003)
- **Yagi Element Diameter Differences Do Make a Difference** (07-01-2000)
- **Notes on Reversible Yagi Arrays** (09-01-2003)
- **I Want to Build a 3-Element Yagi** Parts 1-4 (08-01-1999)
- **Stacking Yagis** 3 Items (09-17-1999)
- **1-2-3: 1 Boom, 2 Bands, 3 Elements Each** (03-07-1998)
- **Three Forward-Stagger 5-Band Yagis from ON4ANT** (08-28-1998, 04-18-1999, 07-16-1999)

Practical Antenna Notes: Upper HF (Mainly): Other HF Arrays and Questions

- **Some Model Quads** 10 Items (04-02-1999, 08-05-2000)
- **New Quad Studies** 10 Items (2000-2002)
- **A Trap Small Quad for 12 and 17 Meters** (09-29-1997)
- **LPDA Design and Modeling Data** 6 items (10-22-1999)
- **Some Notes on LPDA Stubs** (08-10-2003)
- **A 3.5 Octave LPDA of High Potential Performance, But an Antenna that No One Will Build** (04-14-2000)
- **Long-Boom LPDAs for 14-30 MHz** (07-01-2001)
- **3 More 14-30 MHz LPDA Designs** (07-01-2001)
- **4-30 MHz LPDA Design Concepts** (03-01-2002)
- **A 10-Meter LPDA** Parts 1-7 (11-21-1998)
- **Notes on HF General Coverage LPDAs Using 30-35' Booms** (09-01-2003)

- **Notes on a Long-Boom General Coverage LPDA** (11-01-2003)
- **The Birdcage Antenna** (03-18-1997)
- **The L-Antenna** (02-05-1999, 11-25-2002)
- **The B-Antenna** (02-05-1999)
- **17-15-12 and Simple** (05-11-1999)
- **A Vertical Doublet for 30-10 Meters** (12-01-1999)
- **Notes on Hatted Vertical Dipoles for 10 Meters** (01-14-2003)
- **EDZ Beam Update** (05-10-1997)
- **Feeding the EDZ** (08-27-1997)
- **The Expanded Lazy-H** (04-01-1999, 04-15-2003)
- **Curtains for the Extended Lazy-H** (04-15-2003)
- **An Almost Universal HF Back-Up Antenna For the Antenna Farm That Has Everything Else** (02-12-1999)
- **Suppose I Could Have Only One Wire Antenna. . .** (03-21-2003)
- **The Turnstile: An Omni-Directional Horizontally Polarized Antenna** (03-06-2000)
- **Some Notes on Two-Element Horizontal Phased Arrays** Parts 1-4 (05-10-2002)
- **The HB9CV Phased Array and Gain Comparisons** (09-18-1999)
- **Horizontal Phased Arrays with Parasitic Directors** (08-27-1998)
- **Tinker Toys for Adults: Some Notes and Ideas on PVC for Antenna Applications** (08-29-2000)

Practical Antenna Notes: VHF/UHF (Mainly)

- **2 X 3 = 6** (07-25-1997)
- **Experimental Omni-Directional Antennas for 6-Meters** (12-27-2001)
- **The 64-(Euro-)Dollar Question** (01-01-2004)
- **Phased Yagis, EDZ Beams, and Landstorfer-Sacher Yagis** (07-28-1999, 09-16-1999, 04-10-2000, 02-14-2004)
- **The Half-Square on 2 Meters**: Parts 1-3 (04-12-1999, 07-18-2002)
- **Some J-Poles That I Have Known**: Parts 1-4 (01-08-2002)
- **Moxon Rectangles for 6 Meters** (02-03-2003)
- **The Moxon Rectangle on 2 Meters** (10-01-1999)
- **Building a 2-Meter Moxon** (11-01-1999, 12-18-1999)
- **Simplifying the Turnstile Moxon Rectangle Fixed-Position Satellite Antennas** (10-01-2001; 06-01-2003)
- **What's Wrong With This Turnstile Stack?** (05-15-2002)
- **Notes on the Batwing** Parts 1-3 (03-01-2004)
- **In Pursuit of Better VHF Quad Beams: A Work in Progress** (04-21-2001)

- **Notes on Medium-Length 2-Meter Quads and Yagis** (01-24-2003)
- **Circling the Square Quad** (05-01-2004)
- **The Quad Beam as an Amateur Satellite Antenna** (07-01-2004)
- **A Collection of 220-MHz Yagi Designs** : Parts 1-4 (03-29-2001)
- **Multiple Reflectors for Long-Boom Yagis** (10-06-2001)
- **Appreciating DL6WU Wide-Band Long-Boom Yagi Design: Some Preliminary Notes** (11-02-2001)
- **Long-Boom Yagi Sidelobe Suppression** (04-27-2002)
- **Preliminary Studies of Long-Boom Yagis for 420-450 MHz** (04-28-2003)
- **Scaling and Adjusting VHF/UHF Yagis** (12-01-2002, 07-09-2004)
- **An OWA Family of 2-Meter Yagis From 6 to 12 Elements** : Parts 1-2 (08-02-2002)
- **Extending the 2-Meter OWA Family**: Parts 1-3 (09-01-2004)
- **Notes on 6-Element Wide-Band 2-Meter Yagi** (12-26-2002)
- **The OWA Family Moves to 220** (09-02-2002)
- **Wide-Band Utility Yagis for 420-450 MHz**
  - **1. 4- and 6-Element Models** (08-01-2001)
  - **2. An 8-Element Model** (09-01-2001)
- **Some Basics of Very-Wide-Band Yagi Design** : Parts 1-2 (10-01-2002)
- **The Flat-Plane Reflector for 432 MHz** (11-08-2001)
- **Modeling the Double-Diamond for UHF** (12-01-2001, 11-12-2002)
- **Corner Reflectors Revisited**: Parts 1-3 (04-18-1999)
- **Corner Arrays for Personal Communications** (10-01-2001)
- **Planar and Corner Reflectors Revisited** (08-12-2004) (PDF)
- **Modeling the Dual Rhomboid**: Parts 1-3 (08-16-1999)
- **High-Gain, Wide-Band Yagis for 10, 6, and 2 Meters** (09-01-1999)
- **Quagi and Yagi on 2 Meters: Some Preliminary Notes** (04-16-2001)
- **A 3-Moxon Polling Array for 914 MHz** (08-01-2000)
- **The Case of the Curly Collinear** (06-01-2004)
- **A 100-1000 MHz "Utility" LPDA** (12-01-2001)
- **CAP Emergency-Beacon Direction-Finding Antennas** (05-05-2003)
- **Some Notes on FM BC Antennas**: Parts 1-5 (04-10-2003)
- **LPDAs for the 400-800-MHz Television Range**: Parts 1 and 2 (03-20-2003)

Transmission Lines, Impedance Coupling, and Construction

- **12 Ways to See and Love Your Feeders** (05-19-1998)
- **Some (Old) Notes on Home-Brew Parallel Transmission Lines** (01-06-2000, 12-28-2001)

- **Values Along a Transmission Line** (11-25-1998)
- **When is a Quarter Wavelength Not a Quarter Wavelength?** (02-20-1998, 07-15-1999)
- **Some Aspects of Series and Parallel Coaxial Cable Assemblies** (12-26-2001)
- **Some Notes on Modeling Hybrid Transmission Line Stubs** (05-01-2002)
- **Series Matching: A Review** (02-05-1999)
- **Voltage, Current, and Impedance Along a Transmission Line** (04-01-1997)
- **Whose Afraid of a Little Matching?** (04-04-1998)
- **Notes on a Wide-Band 50/75-Ohm Coax Feed System for Low HF Band Dipoles and Vees** (11-29-1997)
- **Link-Coupled Antenna Tuners** (updated 05-20-1999, 12-30-2000, 03-01-2004)
- **Link-Coupled Antenna Tuners: A Tutorial**: Parts 1 - 5 (11-25-1997)
- **Delta, ATUs, and Losses** (03-28-1998)
- **An LCL/LC ATU** (04-01-1996)
- **The Balanced-L Network** (12-13-2002)
- **Rotator Direction Controllers** (01-21-2001)
- **How Accurately Must We Aim a Beam?** (01-22-2001)

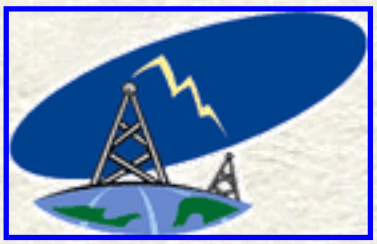
Continuing Series

- **An-Ten-Ten-nas**: a series of articles for *10-10 News* on antenna basics having special relevance to 10-meter operation.
- **Antennas From the Ground Up**: a series of articles on antenna basics for *Low Down*, the journal of the Colorado QRP Club. The emphasis will be largely on lower HF and wire techniques.
- **Antenna Modeling**: a series of articles on antenna modeling specially prepared for *AntenneX*. The series aim is to help antenna modelers get the most benefit and least anguish from their programs.
- **Amateur Radio Continuing Education**: a series of articles prepared over several years for the annual *Proceedings of the ARRL National Education Workshop*. The pieces range in scope from a full-scale proposal for a technical education series down to using a blackboard more effectively in the classroom. There are also links to some other educational sites.

HAMCALC: A Special Note. Numerous articles refer to HAMCALC, a suite of utility calculation programs developed for hams by George Murphy, VE3ERP. The information in the articles gives Murph's address for obtaining the current version on CDROM. Due to a bout of ill-health, HAMCALC is no longer available on CDROM. However, you can obtain **HAMCALC** from this link to the CQ Magazine site and download the current version of the suite in Zipped format. The site provides instructions for installation and use.

-- -- SEARCH -- --

Because the collection of material at the site has grown to fair proportions, I have added a search capability. Clicking on the "Search" button will lead you to a page containing a search engine. You need only enter in a key word (or words) of interest to get nearly instantaneous returns of links to articles containing the term (or terms). Not all articles will be of equal interest relative to your desired term, so use the search engine in conjunction with a scan of the article titles to isolate the items most likely to be useful to you.



Links to Other Antenna Information

These links carry a lot of valuable information and ideas, ranging from antenna fundamentals to advanced topics in antenna design, modeling, feeding, and building. In addition, some provide links to other sites.

Commercial Antenna Manufacturers and Vendors: A collection on known sources, offered because these pages often contain educational as well as commercial information.

Other Amateur Radio Links: A small collection of links to organizations and linkage sites to help you find other good sources of information.

A Final Note

You will note an absence of reviews, analyses, and evaluations of commercially made antennas in the notes at this site. It would be inappropriate for me to remark on such antennas without having the antennas at hand and the appropriate range and equipment for testing them. These notes relate to antenna types and designs over which I have design control and are generally aimed to assist you to understand their operation. Even specific designs are not intended for uncritical replication, although a number of them have been successfully built and used. Still, the goal is not to produce a compendium of antennas for you to build. Rather, the object is to assist you to understand the antennas that you do build, use, or simply think about.

Please do not attempt to download the entire site using software designed for blind downloading. If you wish a record of the entire site, *antenneX* periodically produces a CDROM with the entire site on it. However, I have kept most (but, alas, not all) of the items at the site short enough to read at a single sitting. Pick something of interest, read, and digest. Then pick something else. Let your wandering interests be your guide. If you wish to read more on a subject, by all means, select a related item. Or, look in other good sources for information on the subject. This site is collection of items cast at an intermediate level. Hence, it is far from the last word on any subject, whether you think of basic theory underlying a matter or about very practical construction aspects of an antenna or system.

When you begin to track the items at this site, you become a companion down the path of antenna explorations that I have followed. Be certain that you are ever alert to pathways that are a function of your own interests. This pathway is not the only one, and it is far from the perfect one. But it has been and continues to be both a good and interesting one.

It has been my high pleasure to receive e-mail and regular mail that suggests these materials are of educational and technical service to a broad spectrum of individuals, both in the United States and around the world. Numerous items have appeared in the newsletters and other publications of amateur radio groups. The formal and informal distribution of some of the material, both as written and in translation, in areas where bound publications are unavailable or prohibitively expensive suggests that the energies used to develop and place some of the notes has been productive. So much to learn and so little time to learn it, but always time to share what I have learned along the way--lest it be lost.

Tales & Technicals: the CDROM

As a convenience to those who may pay for on-line time by the minute, have long-distance phone charges, or who have run out of paper downloading items, the site is now available on CD-ROM. The cut-off date for the recently updated Version 4.0 is December 1, 2003. The CD-ROM is available from AntenneX in the "Shopping Shack."

Other information will be made available on these pages as time goes by. For now,

-73-

LB, W4RNL

Updated 09-01-2004

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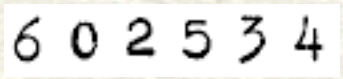

Pages by The House of Two Lions

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13 Sep 2002 II

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9.	TNF400 Instruction Sheet	17 Mar 2003
10	RLB150 Family: Calibration Procedure	25 May 2004

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Amateur Radio Station

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Original, high-quality, radio engineering/modelling programs.

Not freeware -- just free gifts -- please help yourself -- no catch.

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There *ARE* such things as free lunches after all.

Click on Index now.

##

Occasionally programs may be amended. Minor changes such as correction of spelling errors, a more covenient arrangement in the display of results, a change in the number of decimal places, will not result in a change of program name. Only the date of issue will change as shown on a program's introductory screen when a program is run.

When significant changes, improvements or corrections are made in mathematical modelling the program name will be changed and the original program may no longer be available from this site. An announcement of a new name may be made on appropriate Internet newsgroups

and on the "Late-News" page of this website. When a serious error or bug is discovered in a program an explanatory note will appear in the "Program Bugs" page. See Index.

Click on Index now.

Site Index

Navigation: all active text is underlined. Click on the section of interest.

- DOWNLOAD PROGRAMS FROM HERE
- About this Site
- Get Pascal source code from here
- Program Size, Reliability and Accidents
- What kind of programs are they?
- What if . . . . . .?
- Copying and Distribution of programs
- My Portrait
- Back to the Hoarding
- Quotations
- A Tale of Two Old Wives
- What Does the SWR Meter Tell Us?
- Reciprocity - an unpopular conception
- A Bit About Reg, G4FGQ
- Program Bugs
- Feedback from You to Me
- Site Statistics
- LATE NEWS BOARD

Click on "Index" to return here from anywhere on the site.

About this Site

There are few purely decorative features on these pages. No wallpaper, no pictures, decent or otherwise, no voices-off, and absolutely no adverts to distract the serious visitor. Consequently this text and the programs themselves will very quickly download themselves.

The site has been produced by Reg Edwards, UK Amateur Radio Station G4FGQ, for the benefit of the many Radio Amateurs world-wide who make friends and obtain satifaction not only from operating a radio station but also from building and testing it. However, much construction consists of copying or adapting the work of others. But without a quantitative as well as a qualitative appreciation of radio basics the opportunities for innovative work are necessarily severely limited. Refer to Lord Kelvin's pertinent remarks and the great revolutionary work of the self-taught amateur Oliver Heaviside in the "Quotations" page.

Therefore, the primary purpose of this website is to freely distribute computer programs, composed by G4FGQ himself, which mathematically model the behaviour of components, circuits, power amplifiers, transmission lines, aerials, earth electrodes, radio propagation paths, etc. 'Working models' allow experimenters to examine quantitatively how the various input data interact and affect performance. They will relieve the hobbyist of some of the labour, the costs and the disappointments of constructing unsuccessful prototypes. Also of tedious arithmetic. He/she will gain further insight and will better appreciate the finer points of design. Opportunities for original contributions to this truly international hobby will be broadened.

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Download Pascal Source Code from here

For various (some dishonest) reasons I have been asked on numerous occasions to release the source code of programs available free from this website. To satisfy the genuine curiosity of those people who have asked for "the 'formula' for .... ", here are a few source code examples. Anyone who considers he understands the contents of these *.pas files is hereby granted permission to make full use of the information contained therein. If any future products result no reference shall be made to the author who is unable to provide any assistance in such matters.

Files are best read in non-proportionally-spaced text such as displayed by the mini word processor "NotePad".

To download a file click on underlined program name.

GRNDWAV3.pas * Groundwave propagation vs frequency, distance and terrain.

TOPHAT2.pas * Performance of top-capacitance loaded vertical.

PADMATCH.pas * T and Pi resistive-matching and minimum loss pads.

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Program Size, Reliability and Accidents

The number of possible ways in which a computer program can misbehave increases enormously with its size in bytes. A very large program is virtually certain to contain defects or bugs if only because of the impossibility of identifying and testing every one of these many ways. Debugging operations themselves are likely to introduce further defects. The importance of individual defects is uniformly distributed randomly between trivial and catastrophic.

A smaller program is a more reliable one. Unnecessary elaboration must be removed. Simplicity of source-code structure and understandability is the ultimate target. Small programs are better understood by both the programmer and during later maintenance. Testing will be more thorough and 'bugs' far less likely to remain. Consequently large programs must be composed of a number of small, independent, testable modules whose functions are precisely, unambiguously defined beforehand. Each module is then itself a complete small program.

The size of a complete program can greatly increase due to built-in precautions to protect itself, the computer and its user from the consequences of accidental entry of 'illegal' data. For example, illegal data might result in the maths processor being asked to calculate the square root of a negative quantity. To omit protection might result in the program 'crashing', sieze-up of the computer's operating system and the need to press the re-set button. The number of possible ways in which illegal data can be entered is also very large. To lay traps, deal safely with them all and inform the user, is impractical and will itself increase the liklihood of bugs

remaining in the finished product.

In G4FGQ's programs a bare minimum of protection is provided against entry of impossible data. Inevitably the user will make mistakes especially when unfamiliar with a new program. Mistakes are most likely to occur when entering numerical values. They fall into categories of incorrect measurement units, impossible physical proportions, ie, wire diameter greater than wire pitch, inclusion of non-numerical characters and using a comma for a decimal point. Obvious errors can be corrected, of course, by inspecting the data before pressing the Enter or (cr) key.

If you are running one of these programs in DOS where you have arrived from Windows, and it aborts, the system may not automatically switch you back into Windows. To return to Windows from DOS type EXIT (cr) against the DOS prompt. No harm is likely to come either to the computer or to the program as a result of being unceremoniously dumped back into the system.

To re-start a program which has just aborted while in DOS, from the same directory as the program is filed, just type the program's name against the DOS prompt followed by (cr). The .exe ending may be omitted. If you are not in the same directory then the name of the program must be preceded by the directory path to the program, the several items being separated by backslashes. Or you could change the current directory to that where the program is installed and then enter the program's name.

To Q(uit) or E(xit) from within a program do so by pressing the key indicated on the menu bar (Q or E) at the bottom of the screen. The option to quit is available only when the menu bar is visible. Pressing the quit key when the option is not available, if followed by the (cr) key, may indeed cause the program to close down but not in a very graceful manner.

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What kind of programs are they?

The source-code language of G4FGQ programs is Turbo Pascal. Unlike BASIC and various spreadsheet calculators, when compiled into a stand-alone program, as they all are, the Pascal source text is inaccessible and cannot be modified. Programs can be run only as the author intended. Integrity and authenticity are preserved.

Subject matters are technical and are necessarily dealt with quantitatively. At the heart of each program is a set of mathematical functions and procedures. Properties of the 'parts' used to construct the model are stored within the program. Other procedures accept from the experimenter the details specifying the model. The program then proceeds with its analysis and its conclusions are displayed. Advice or warnings may be given to the experimenter as appropriate.

Entry of numerical input data obliges the experimenter to clearly visualise the system modelled and approach it logically. He will learn to ask himself the right questions. The relative importance of the various factors affecting behaviour of the system will be better appreciated. This conforms with one of the purposes of amateur radio activities: self-teaching in the art of radio communication.

But before the amateur who takes pride in being purely practical and ignorant of theoretical matters now leaves, don't go, these programs have been written with you in mind.

To use these programs you need no more than the ability to estimate the number of rolls of wallpaper required to decorate your shack. That's not maths - it's only arithmetic. Authors of articles published in radio magazines often confuse these activities.

Programs are intended to relieve the user of the labour and tedium not only of mathematics but of arithmetic too. But the quantitative visualisation of a model cannot be avoided.

If you have not yet mastered the sizes of enamelled copper wire in terms of milli-metres and still work in 'thou of an inch', or 'mils' you will be at a slight disadvantage - all the programs are in metric units. But down-load a program anyway and let me know how you get on with it. You'll be back.

Actually, most of my programs are more than 10 years old. Most of the basic research had been done and the mathematical procedures had been developed by then. Recent work has been to extend their range of application, to combine related subjects into single programs, to generally tidy up loose ends, and to make programs presentable and user-friendly.

At present, several more are in their final testing stages, being cross-checked for mathematical inconsistencies, bugs, etc. More effort is spent on later quality-control than on producing the first useable version of a program.

It's good practice to put a supposedly finished program on the shelf for a month or so, allow one's mind to forget it, concentrate on something else such as producing this web site, even spending some time on the 'bands'. It is surprising how previously unnoticed defects then become obvious. However, these discoveries do not improve one's self-confidence - how many other hidden boobs remain?

Until recently I have never actually finished a computer program. The self discipline of saying "It's done. Date it. Archive it. Make it available on the Net" has not come easily.

Index

What if . . . . ?

All programs ask for a set of numerical values which describe or define the matter in hand. Often the data are physical dimensions plus a choice of operating conditions. This initial array of input data is always on view at the top of the screen.

When a set of results has been computed it is displayed in the lower part of the screen. It is now a simple matter to select and change any one of the initial values while asking "What if I set it to this particular value . . . ?" Immediately the question is answered. A similar question can then be asked about any other input parameter.

Later programs, on the press of a key, have the facility to sweep input parameters slowly in fine increments over wide ranges while immediately observing the effects on output data.

There is no need to go back to the begining and re-enter a complete set of initial values although that particular option is available when needed. The response or behaviour of complicated systems when subjected to changes in inputs can be analysed and appreciated very quickly. Not only will programs be of practical use - by encouraging the user to become familiar with magnitudes and ranges of the essential factors they will also be of educational value.

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Copying and Re-distribution of Software Programs available from this Website

There are no restrictions on copying and re-distribution of self-produced software programs available free from this website, via any electronic, magnetic or optical means, by any individual or organisation, except that no copies shall be made, distributed or sold for commercial or financial gain.

With due regard to the foregoing I retain the Copy and Intellectual Property Rights for all programs. To preserve program integrity the source code of these programs is not normally available.

Signed:

R.J.EDWARDS

Radio Station G4FGQ & Website Proprietor

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Amateur Radio Station G4FGQ

Reg

"You see, I'm not an ogre. I'm really a kind old chap."

Return to Index

Send comments, queries, complaints, requests, suggestions, ideas to Reg, G4FGQ

NYLO'S NOTEPAD

Homebrew projects for radio & electronics enthusiasts

- ANTENNA PAGE -

back to HOME page

RAISING WIRE ANTENNAS

Wire antennas deliver the most bang for the buck, but they require a little work to put up. The most common way to put one up is to use the biological towers in your yard: the trees! To make this work, you need to install ropes to hold up the antennas. Here's a description of my favorite technique for getting those ropes up there:

SETTING ROPES IN TREES FOR WIRE ANTENNAS

ANTENNA PROJECT SELECTOR

VHF ANTENNAS	
2 meter HT Handipole	Add some gain over your 'rubber dummy load' with this full size dipole on your HT.
2 meter rollup J-pole	Compact portable companion for travel or temporary operation. Hang it up anywhere.
2 meter 4 element, portable yagi	This yagi breaks right down and the elements store right inside the boom. The boom and mast can be used for walking sticks. Can be used as a base, as a portable or for foxhunting.
2 meter, 4 element quad	Rugged, yet lightweight powerhouse is great for base or portable and is especially *excellent* for foxhunting. It can also be constructed to break down for compact storage.

2 meter Bicycle whip antenna	This 1/2 wave, end-fed vertical with compact feed is built right onto a 1/4" fiberglass bicycle flag whip and *requires no ground plane nor radials*!. Excellent for bikes, boats, RV's, etc., or anywhere a compact vertical is desired.
MONOBAND HF ANTENNAS	
10 meter wire J-pole	This end-fed half wave radiator is simple to make and even simpler to put up with a single rope. Can also be strung horizontally from your shack's window with a single rope to another support.
10 meter wire collinear 'super' J-pole	Yeah! More gain! Two 5/8 wave, phased radiators. Can be strung vertically for omnidirectional gain. This is a great omnidirectional gain antenna!
20 meter wire J-pole	This end-fed half wave radiator is simple to make and even simpler to put up with a single rope. Can also be strung horizontally from your shack's window with a single rope to another support.
40 meter wire J-pole	UNDER DEVELOPMENT
MULTIBAND HF ANTENNAS	
80-10 meter doublet with open wire feed	Arguably the most popular 80-10 multiband antenna ever! Certainly, the most bang for the buck!. If you only build one, this should be it. Ultra simple to build and no pruning required. Good lobes of serious gain on the higher bands and low loss. Requires an outboard tuner with open wire terminals for best performance or you can use your rig's built-in tuner and a remote balun to match most bands. This antenna has the most gain when strung as a horizontal, flat-top, but it is versatile and can be strung as a sloper, or inverted vee.
80-10 meter mobile vertical	My latest project, and the results of 9 months of experimenting after putting an HF rig in my car. The signal reports have been excellent! This is, essentially, a homebrew version of a Texas Bugcatcher, except that the capacity hat is up at the top of the whip where it belongs for maximum efficiency. And, it's a lot cheaper!

Good luck and have fun with your antenna project!

--...MARK_N1LO...--

These antenna plan files are in Adobe Acrobat Portable Document Format (*.PDF files), the emerging standard for downloadable documents of all types on the web. To read or print the item, just click on it to download the file, then open it using the free Adobe Acrobat reader program, which you must have installed on your computer. Typically, Acrobat will open automatically and display your goodie. Follow the link below to get Acrobat if you don't already have it.

Here's where you can

Have you found something useful here? Got a comment? Drop me a line: n1lo (remove 'nospam' from address when sending)

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.

THE N4UJW ANTENNA DESIGN LAB

THE BEST HAM RADIO OPERATORS HAVE GOOD ANTENNAS!
YOU CAN'T WORK THEM IF YOU CAN'T HEAR THEM!

DESIGN OR BUILD YOUR ANTENNA HERE!

Antenna project contributions from all over the world!
HF, VHF, UHF AND LOTS MORE!
All projects have been offered by their authors
to share will all hams worldwide!
A note of caution, we've re-calibrated our Humor Filter.
It still has a few bugs, but is working better now but use caution!
Please feel free to take a look around now at these most popular
antenna projects and information and come back later
when the antenna farm is in full bloom!
Some of your clicks below will take you away from our site,
so click your back button to return here.

THE 2 METER SSB SQUARE LOOP
NEW 6 METER ROTATABLE DELTA LOOP
HOW TO USE AN ANTENNA TUNER
2 METER HENTENNA
BUILD AN ALL BAND "UGLY" BALUN
TWO CLASSICS THE DIPOLE AND THE INVERTED V
6 METER ZIP CORD EDZ 50 OHM FEED
BUILD AN ALL BAND STURBA CURTAIN!
ANTENNA PROGRAMS DE VE3SQB YOU MUST SEE!
BASIC YAGI DESIGN FOR HOMEBREW
2 METER SSB YAGI ANTENNA PROJECT
MULTIBAND HF DIPOLE PROJECT
2 METER SLINGSHOT RE-VISITED
6 METER HALO LOOP
6 METER EDZ SIGNAL SQUIRTER
3 ELEMENT BEAM FOR 6
ANOTHER 6 METER LOOP
THE ONE ELEMENT BEAM! (Also see next listing)
ONE ELEMENT BEAM 20 THRU 6! Fantastic!
THE CLOUD WARMER NVIS BEAM
THE HENTENNA RE-VISITED
DESIGN A J POLE FOR ANY BAND
40 METER HUSTLER MOBILE ANTENNA MATCH PROJECT
EASY 2 METER GROUND PLANE +
DOUBLE BAZOOKA! SET YOUR SIGHTS ON BETTER QSO'S
THE FOUR DOLLAR SPECIAL (save those bucks)
THE LATTIN 5 BANDER

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DECIMAL CONVERSION CHART FOR ANTENNA BUILDING
THE WA7CS TOWER PLANTING PROJECT
160 THRU 2 METER ANTENNA MARVEL! YA GOTTA SEE THIS!
BUILD A 2 METER 5/4 WAVE ANTENNA
ANOTHER GREAT J POLE DESIGN FOR 6 METERS
TRY OUT THAT NEW 40 METER ANTENNA HERE
VISIT THE CUSHCRAFT ANTENNA TECHNICAL LIBRARY
THE "LL COOL J" - 2 METERS FOR A SONG
20 METER "HAMSTICK YAGI"
20 METER 2 ELEMENT HAMSTICK VEE
GET A GOOD GROUND! Read the article!
NEW COAX-FED SLOPER LONG WIRE BY LA1IC
NEW SLOPING INVERTED V BY LA1IC
NEW 3 HALF WAVELENGTH VEE
NEW INVERTED VEE *LONG WIRE*
MORE ANTENNA SITES TO VISIT
K5AT ANTENNA LINKS
ANTENNAS "THE EASY WAY" DISPELL THOSE MYTHS!
ANTENNA THEORY AND TRAINING
PRESENT BY THE UNITED STATES NAVY
(Some files below are under construction)
WAVE PROPAGATION

Do you have a great antenna project or antenna building experience you would like to share with your ham friends worldwide and would like it published on HAMUNIVERSE.COM?
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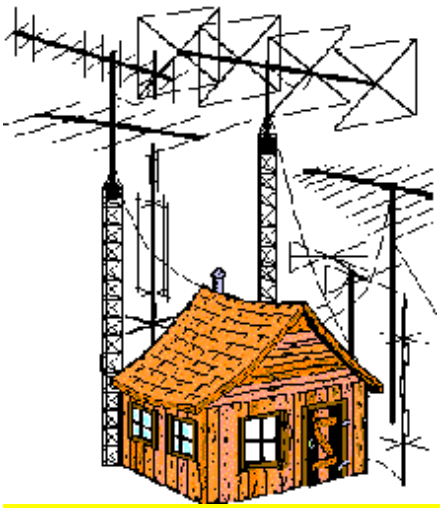

USE THE BARNS AND NOBLE LINK BELOW TO FIND THAT ANTENNA DESIGN BOOK YOU'VE BEEN WANTING! FANTASTIC SELECTION.. USE THE WORDS "HAM ANTENNAS" OR "ANTENNAS" IN THEIR SEARCH WINDOW TO SEARCH THEIR LIST.

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Get those questions answered or add your input!
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SITE MAP

Webmasters note:
I Apologize for this very cluttered page!
This web site has grown so fast that I cannot keep it mowed. The weeds are getting very tall in places and the grass needs cutting just about every other day!
Broken antenna elements are scattered all over, bent towers, burned out baluns, loose connectors and lots of spiders, useless bits of wire of all sizes and the accumulation of SWR is getting awful!
Please be patient.....
We plan to use weed, grass killer and insect spray so the page should be much neater in the future! **73** N4UJW.

NEW KE0VH DIGITAL PROJECT

COAX DATA

The Wonderbar Antenna

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73 N4UJW HAMUNIVERSE.COM

SITE MAP

THE N3SDO "J" BEAM!

Share your homebrew antenna project!
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Book Home Page

ARRL Antenna Book

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Antenna Magic

by

Ray Jurgens

KQ6RH

(C) 1998, 1999, 2000

Updated 2/25/2000

This page is devoted to light weight wire beams and quad antennas. Most of these antennas are low profile wire beams constructed with fiberglass rods and tubes supported with central hubs. Each of these has been designed to give significant gain over a dipole but require much less space. All are self supported and able to be rotated with low cost TV antenna rotators. This site provides construction details as well as information about where specific materials can be found. We also offer some materials that are difficult to find or make.

Philosophy

You may wonder what the motivation is to try to build antennas with super-light weight structures when larger and heavier structures are readily available. Some might say that this is equivalent to buying into the minimalist movement which, for ham radio, is the same faction as the QRP movement. However, there are many reasons for developing light weight systems for portable and emergency situations as well as relatively stealth applications in controlled spaces such as apartment and town house developments. A few general principles that we have tried to abide by are that each antenna must be self supporting, must be sufficiently light in weight that a single person can easily assemble it and erect it, and that if at all possible, it should be able to give significant gain over a simple dipole. Why settle for less? This is not exactly minimalism, this more in the line of optimism.

Content

This page currently discusses quad-loops, cubic quads, Pfeiffer quads, VK2ABQ wire beams, Reflected M beams, Half Squares, X-Beams and a few simple antennas including a rotatable dipole. The Reflected M beam is a variant of both the VK2ABQ beam and the X-Beam. For a given wavelength, the Reflected M beam is the smallest of all. To compare these, I'll consider the length of the spreader required for a 15 meter antenna cut for 21.2 MHz. Though these differences are small, they are significant in that the bending of the spreader increases as length cubed for an applied force. Thus, the bending of the Reflected M spreader is reduced to 60% of that of a quad loop for a similar wire loading. The Reflected M beam is planar, takes less space to construct and can be designed with performance nearly equal to that of the two element quad. For example, a two element optimized quad might have about 7.5 dBi gain with a F/B ratio of 17 dB. The optimized Reflected M has 6.0 dBi gain with an F/B ratio of near 20 dB, though the actual rear lobes are about 2 dB poorer. The reflected M beam can be assembled and erected in about a half hour. Try that with any quad. I also give the performance of the X-Beam which is similar to and has similar performance as the Reflected M. Though the X-Beam has a good front to back ratio, the front to side lobe ratio is much poorer than that of the Reflected M. Still, the X-Beam is a great antenna that uses aluminum tubing for the radial elements and folded back wires to complete the element lengths rather than using fiberglass and wire, but the Reflected M is better in most respects.

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Comparing Antennas

The table below give some idea of how each antenna compares to the others. The sizes given are for a 15 meter version of each antenna. The size reported is the required length of the spreader. The impedance of some antennas depends upon several parameters, but the typical values are given below for both free space and ground mounted cases. In the case of the cubic quad, only the two element version is considered in the table. The elevation and peak gain reported are for the first or lowest lobe above an flat earth ground with average properties. For the elevation data, all antennas have their centers set at one half wave length above the ground or about 24.5 feet except for the Quick Vertical which is mounted 3 feet above the ground. The Quick Vertical uses a Hustler whip fed against a counterpoise. The Half Square antenna is actually better at suppressing high angle radiation if it is set at about 18 feet above the ground, but the data given in Table 1 is for 24.5 ft.

Antenna Type	Size ft	Z Ohms	Gain dBi fspc	F/B ratio	Elev. degrees	Peak Gain dB
Quad Loop	8.38	130, 125	3.2	0 dB (bi-dir)	24	7.8
Cubic Quad	8.38	110, 117	7.4	17.7	24	11.4

VK2ABQ beam	8.45	146, 200	3.8	12.6	26	9.8
Reflected M	7.46	38, 47	4.7	11.8	26	9.8
Reflected M	7.40	22, 24	5.8	21.8	26	10.6
Half Square	11.8	48, 50	4.4	0 dB (bi-dir)	14	3.3
Quick Vertical	8.0	24, 34	0.7	0 dB (omni-dir)	24	-0.9
X-Beam	9.3	34, 33	5.7	22.9 (deceptive)	26	10.5
Dipole	11.2	71, 65	2.1	0dB (bi-dir)	26	7.6

Table 1

Properties of Various Antennas in Free Space and Elevated a Half Wavelength

Inspection of the table gives some interesting insights:

1. The Reflected M beam has the smaller turning radius and size
2. The only antennas with good matches to 50 Ohm line are the Reflected M and the Half Square
3. The Cubic Quad has the highest free space gain followed by the Reflected M and X-Beam
4. The Reflected M and the Cubic Quad have the best F/B ratios
5. The Half Square wins the low launch angle by a large margin
6. The Cubic Quad has the highest ground mounted gain followed closely by the Reflected M and X-Beam

All of these antennas have respectable performance, and all can be designed with light weight fiberglass construction. So, take a look at these, and pick a winner for your application.

Main Menu

VK2ABQ wire beams
Reflected M beams
Quad Loops
Light Weight Cubic Quads
Pfeiffer Quads
Half Square

Quickie Vertical
X-Beams
Dipoles
Construction Hints
Properties of Fiberglass Rods and Tubes
Our Products
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KQ6RH Home Page

Return to Main Menu

LINKS

Check Out KQ6RH's home page. This page contains much information on using

Hustler Whips in various configurations for use in mobile, apartments and town houses.

KQ6RH HOME PAGE

Go directly to Max Gain Systems to check out Fiberglass rods and tubes.

MAX Gain Systems

Check out the Northern California DX Foundations synchronized beacons.

NORTHERN CALIFORNIA DX FOUNDATION

Check out the Ionospheric propagation; the following links are useful!

NOAA Space Weather Links

NOAA Geomagnetic Indicies

NTIA HF Propagation Models

NWRA Real Time Sunspot Number

NWRA IONO Predicts

NWRA World Wide Ionosonde foF2

NRL Beacon Scan

AM LOOP ANTENNAS

MOVED ---

--- to its own semi-domain at: http://www.mindspring.com/~loop_antenna. This will hopefully solve the bandwidth problems that shut it down at its former addess. Sorry for the inconvenience and re-direction.

QRZ.COM

Build them again!
MAIN INDEX 7 Sep 2004 @ 01:43:02 UTC

Antenna Programs

Back to the main file index

Name	Size	Date	Description
40mbeam.zip	6964	02-02-1998	Instr. to build a beam for 40 m - K5DKZ
antdl6wu.zip	19040	02-02-1998	Yagi Design for 50 MHz and up
antenna.zip	135028	02-02-1998	UHF/SHF Antenna design
antennas.zip	86768	02-02-1998	Antenna design article collection - KB4YLY
antfo.zip	17237	02-02-1998	Yagi Design prg for 144 MHz and up - WA2TIF
ariel15.zip	32705	02-02-1998	HF wire ant & trans line design v1.5 - WB4YZA
ariel30.zip	27938	06-24-1997	HF wire antenna design program by WB4ZYA
asa11.zip	210312	02-02-1998	Antenna system Analysis v1.1 for Windows
azprj104.zip	1102740	06-24-1997	Postscript files generate az. equidistant
bazooka.zip	11025	07-07-1997	Calc Double bazooka - Win3x/Win95 - W4BEJ
coax1.zip	21630	07-07-1997	Coax calculator for Windows
coaxtrap.zip	578380	08-05-1997	Calc coaxial traps (Win95/NT) - VE6YP
dipole.jpg	80898	01-23-2002	Easy dipole design by AF4NB
discone1.zip	5758	02-02-1998	Discone Antenna Design - K5DKZ
esprop14.zip	146495	02-02-1998	ES-PROP v1.4 Sporadic-E prop analysis prog
hamftz03.zip	48669	02-02-1998	Ionospheric prop prediction program by HB
helix_20.zip	31495	02-02-1998	Helix antenna design - OH0NC
jpole.jpg	49847	04-08-1999	J-pole Design picture
jpole.zip	20052	02-02-1998	J-pole Design Prg V1.1 by WA2ISE
jpole1.zip	15354	02-02-1998	J-pole Design picture - BMP by KI7ZD
logyag16.zip	32083	02-02-1998	Yagi with Log-Cell feeding design - OH0NC
loopcalc.zip	10945	09-19-2001	LoopCalc 1.3 calculate magnetic loop antenna
lpda.zip	7488	02-02-1998	Excel sheet to design Logperiodic Dipoles
mast.zip	33514	02-02-1998	Calculate mast dim. needed v1.0beta - KD4NUE
mloop.zip	8864	02-02-1998	Miniloop 1.1 - Small signle turn ant design
mloop31.zip	401268	08-05-1997	Magnetic Loop Ant Design v3.1 - DK1NB

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msdsp034.zip	68563	08-13-1997	Meteor Shower Receive
mssof42b.zip	273222	02-02-1998	Meter-scatter v4.2b - predicts meteor showe
mssof42e.zip	274499	02-02-1998	Meteor-scatter v4.2e - predicts ope
mssoft43.zip	584396	08-13-1997	Meteor Shower transmit and prediction - OH5IY
necfpc.zip	248018	02-02-1998	NEC2 w/32 bit reals (for Powerstations)
rfprop.zip	65575	06-24-1997	Radio Prop calcu. v1.01 for Win3x
stress.arc	15151	02-02-1998	Mechanical design for antennas
tl.zip	31612	02-02-1998	TL Transmatch Calculations v1.7 - N6BV
trap01.zip	4964	02-02-1998	40/80m trap dipole construction - K5DKZ
voawin.zip	5757997	06-24-1997	Estimate freq coverage hour/month (Win ver)
wndipole.zip	4321	02-02-1998	Dipole calculations for Windows - N4PVU
ya101pat.zip	84786	02-02-1998	Yagi analyzer patch for YA v1.0
yagim311.zip	693013	02-02-1998	Yagimax 3.11

Main File Index

Return to the antennas file section.

Welcome to

AA3PX's Antenna Page

Dedicated to Harold H. Beverage, other radio Pioneers and fellow enthusiasts with a BIG yard.

My 160 meter vertical.. 80 Ft Shunt feed, top loaded.

Harold H. Beverage PAGE.....(many photos, maybe slow loading)

Beverage Antenna Design...(If you have the space for it)

Inverted"L" info

AC6V's Great Ham Page

G5RV antenna spec's and info

Nordic Shortwave Center

AA3SJ's QRP Page

Tip's and Tricks Antenna Construction Info from readers !

(Please E-MAIL me with your Construction Tips and Tricks for posting!)

Interesting places to see

AM broadcast history site (very nice)

Longwave page

Nikola Tesla page , Check this out !!!!!

Weather info Observed at

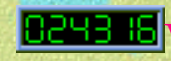
Excellent NOAA Weather Site

Satellite images

If you have any questions, comments, would like to add a link to this page, or help. Please

contact me :

aa3px at localnet dot com



VISITOR'S

PLEASE CHECK BACK . Any input would be greatly appreciated.

QSL.net

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[All Lists]

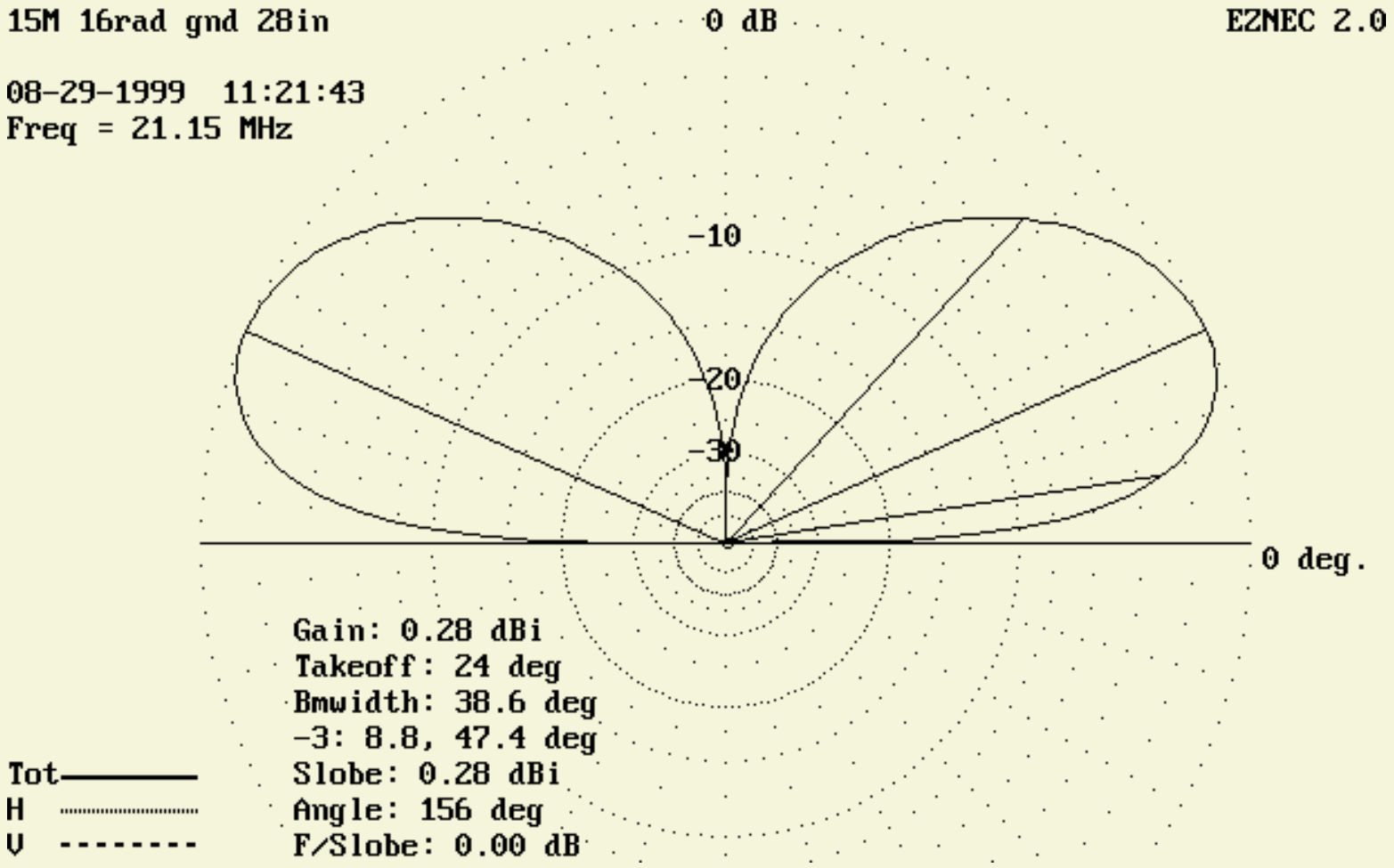
Common Antennas For Portable Use

Lots of speculation goes into antennas for portable use. There are conflicts. On the one hand you need to cart all the stuff with you so there is a definite "less is more" attitude here. On the other hand, putting up a "dummy load" does no one any good. As it turns out, height of the pole or poles available in wavelength determines just about everything in the selection process.

Several basic types are compared here using EZnec software. One is a very highly optimized vertical. It uses 16 elevated radials 28 inches off the ground. It uses lots of segments and is carefully matched to the EZnec model to make as certain as possible that it is free of computational artifacts.

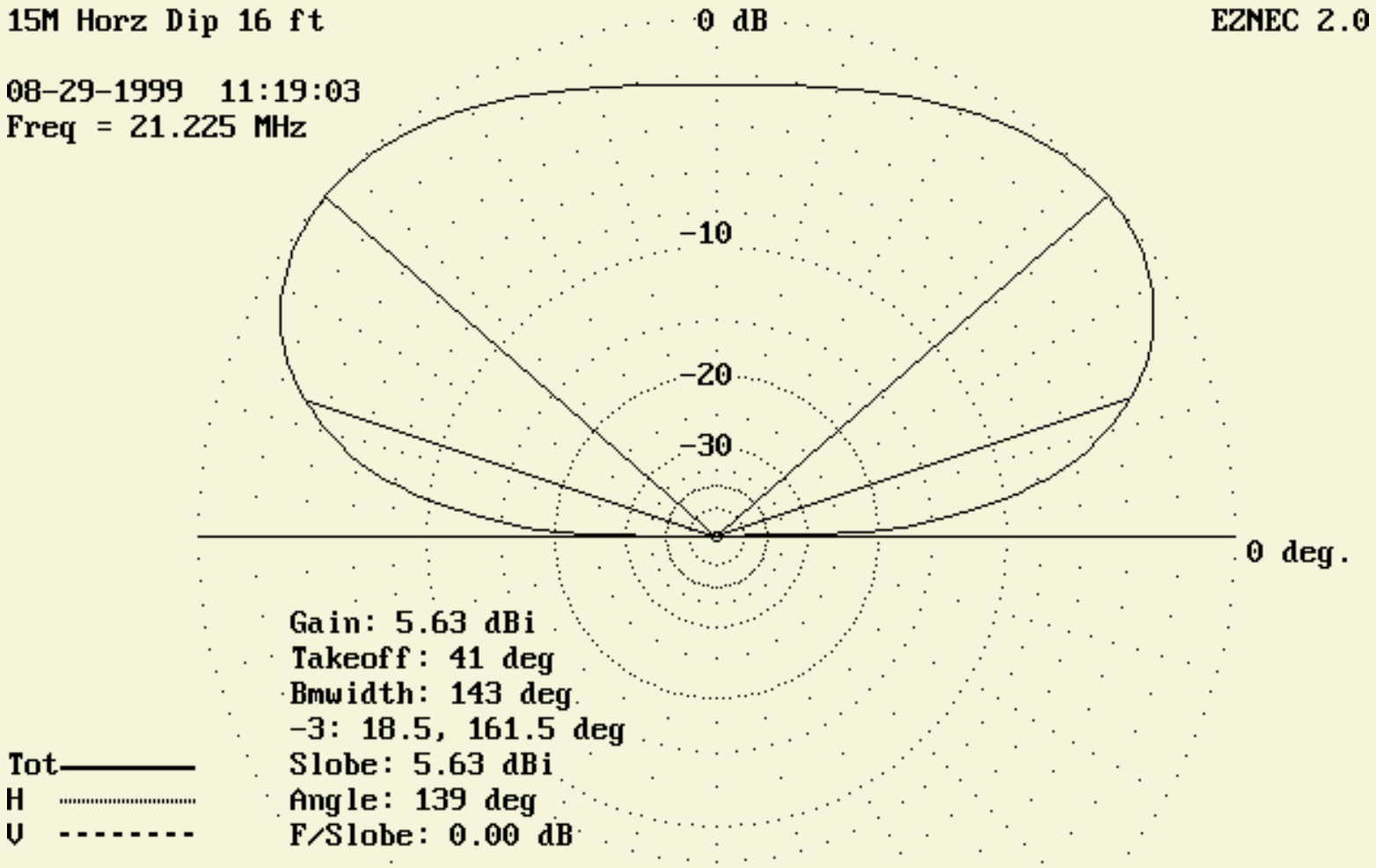
The second is a classic dipole erected using about the maximum length of pole that can be easily packed or taken on an airplane, a pair of multi-segment 16 foot poles. The third is a classic inverted V with an apex angle of 90 degrees. Another was modeled with an apex angle of 120 degrees but it had a large footprint due to the long guy wires. It was a tiny bit better at the flatter angle. The final one was an interesting fan dipole with triangular wire elements that were 24 inches wide at the far end. All used the same pole height.

The shock was the vertical. First look at the great pattern!:



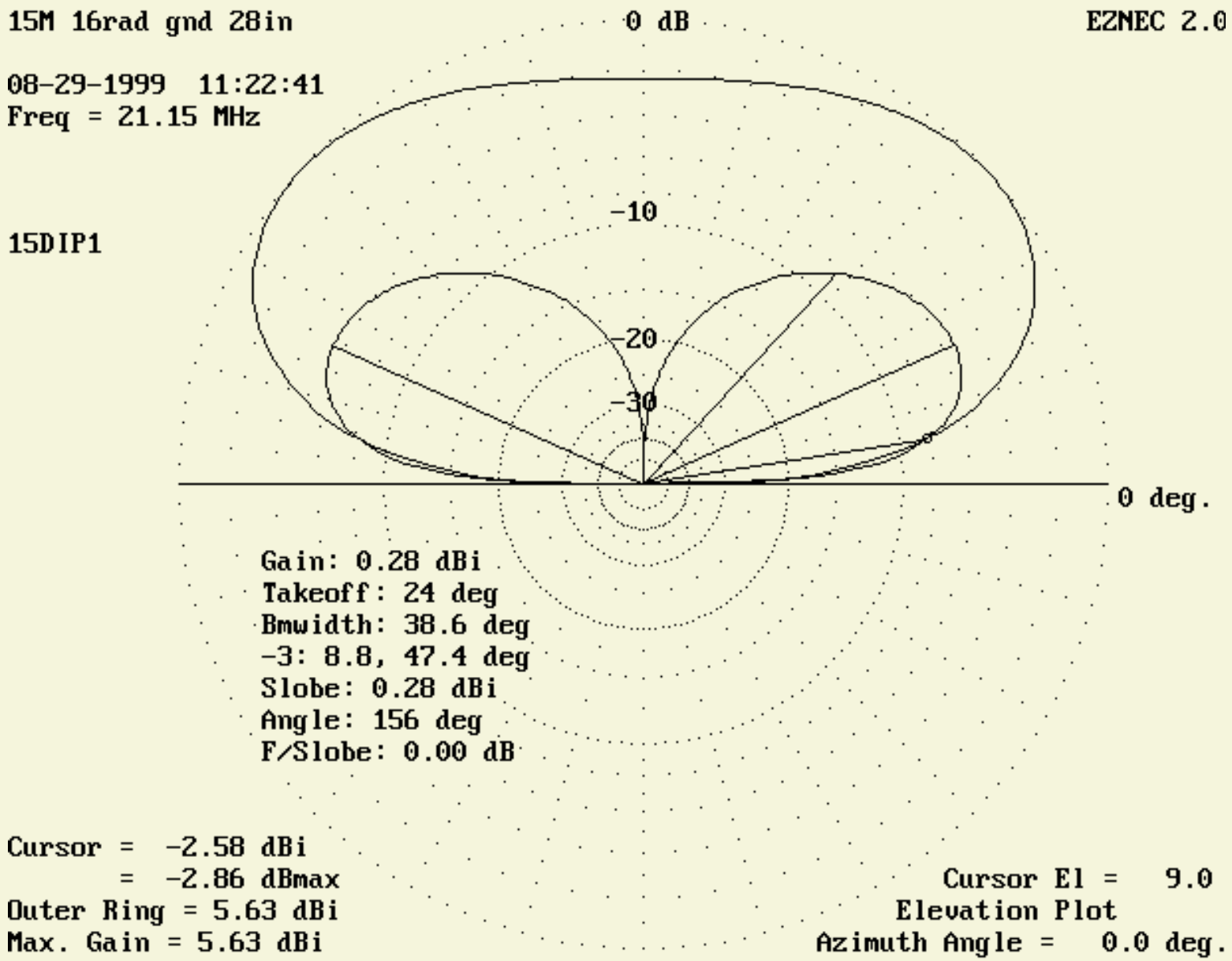
This looks **REALLY** good. It should. Everything that can be done has been done. This design brings along 16 additional stakes each about 3 feet long, to hold up the 16 elevated radials, but it is self supporting and needs no guy wires other than those on the 16 stakes for the elevated radials. Look at the nice concentration of power in the low angles, something a vertical is famous for. This one is getting close to the pattern over perfect ground.

Let's look at the pattern of the dipole. Who in their right mind would cart along two 16 foot poles instead of the above antenna. You could even get away with not elevating the above radials and laying them out on the ground as long as you used 16 or more of them, and still get much the same performance as above. But here is the dipole pattern:



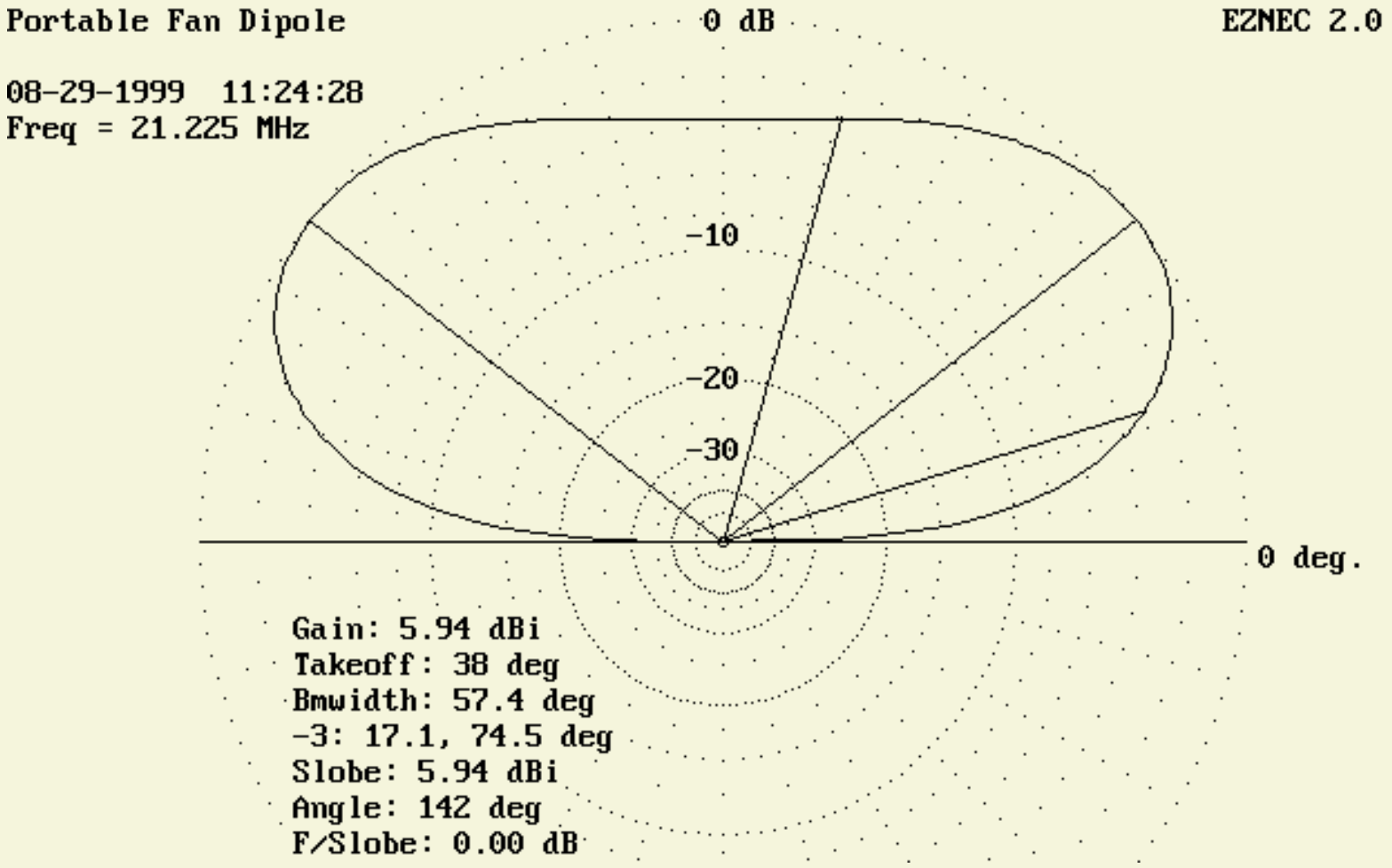
Not much to look at compared to the beautiful vertical pattern above. But there is something to look very very carefully at here. The absolute values. All of these antennas were modeled on 5 millisiemen ground and in each case the support poles were modeled as well as grounded metal poles. Note that the presence of the poles made only a slight difference in the patterns and gain figures. All antennas were carefully resonated at the test frequency.

So what is wrong? Let's have a look at the two patterns scaled and plotted on top of each other:



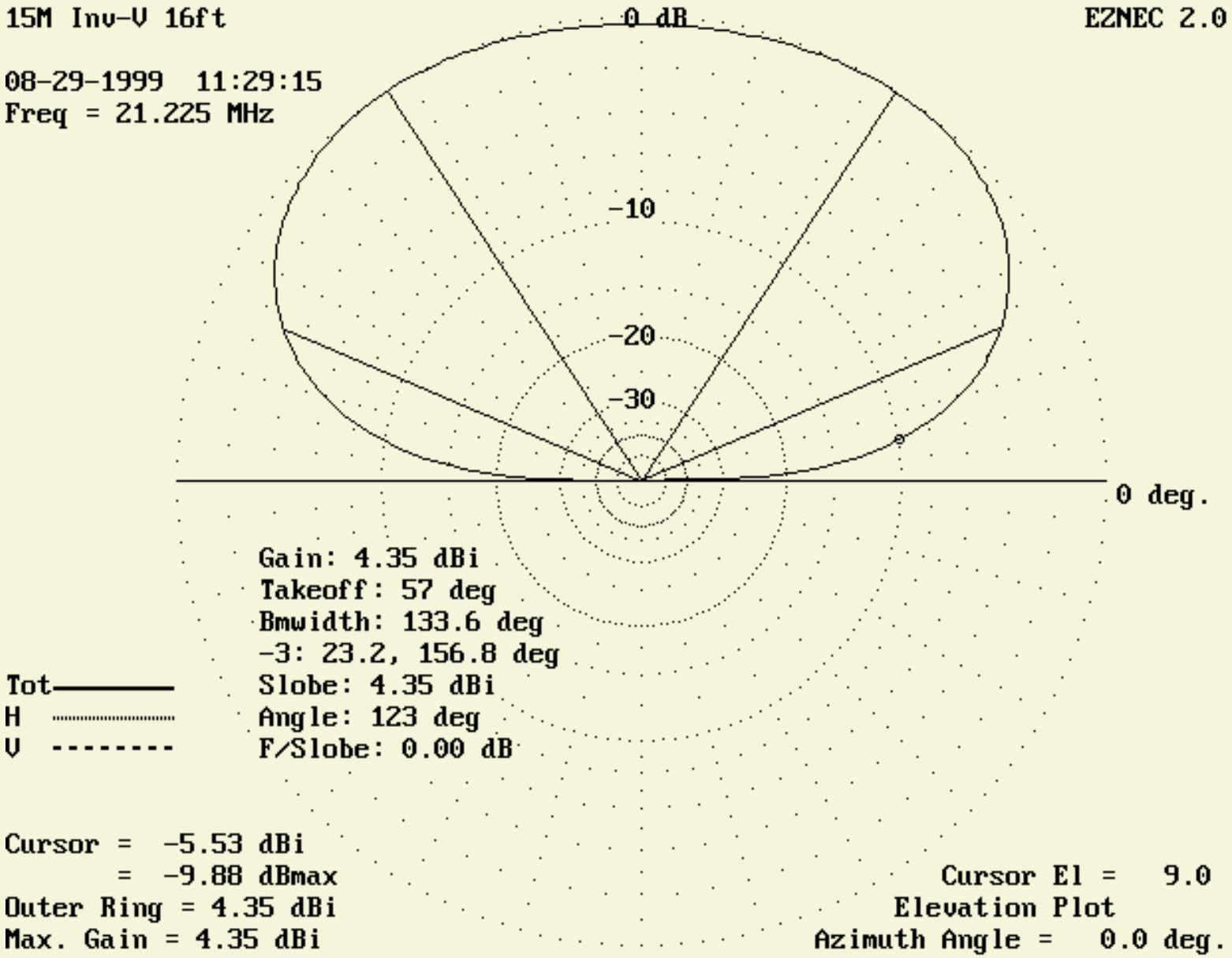
Oops. Suddenly the dipole does not look so bad. In fact, there is only a very tiny portion of the patterns, below 9 degrees, where the vertical is better than the dipole and there only by the smallest of fractions of a db. The vertical pattern is a subset of the dipole pattern. At angles like 15 degrees, a solid DX angle, the dipole is considerably better. And this dipole is only mounted about 3/8th wave above the ground.

The fan dipole was fun. It was of course shorter between poles than the regular dipole and seemed to have a slightly better gain figure vs. the dipole. Perhaps it had a bit bigger capture area:

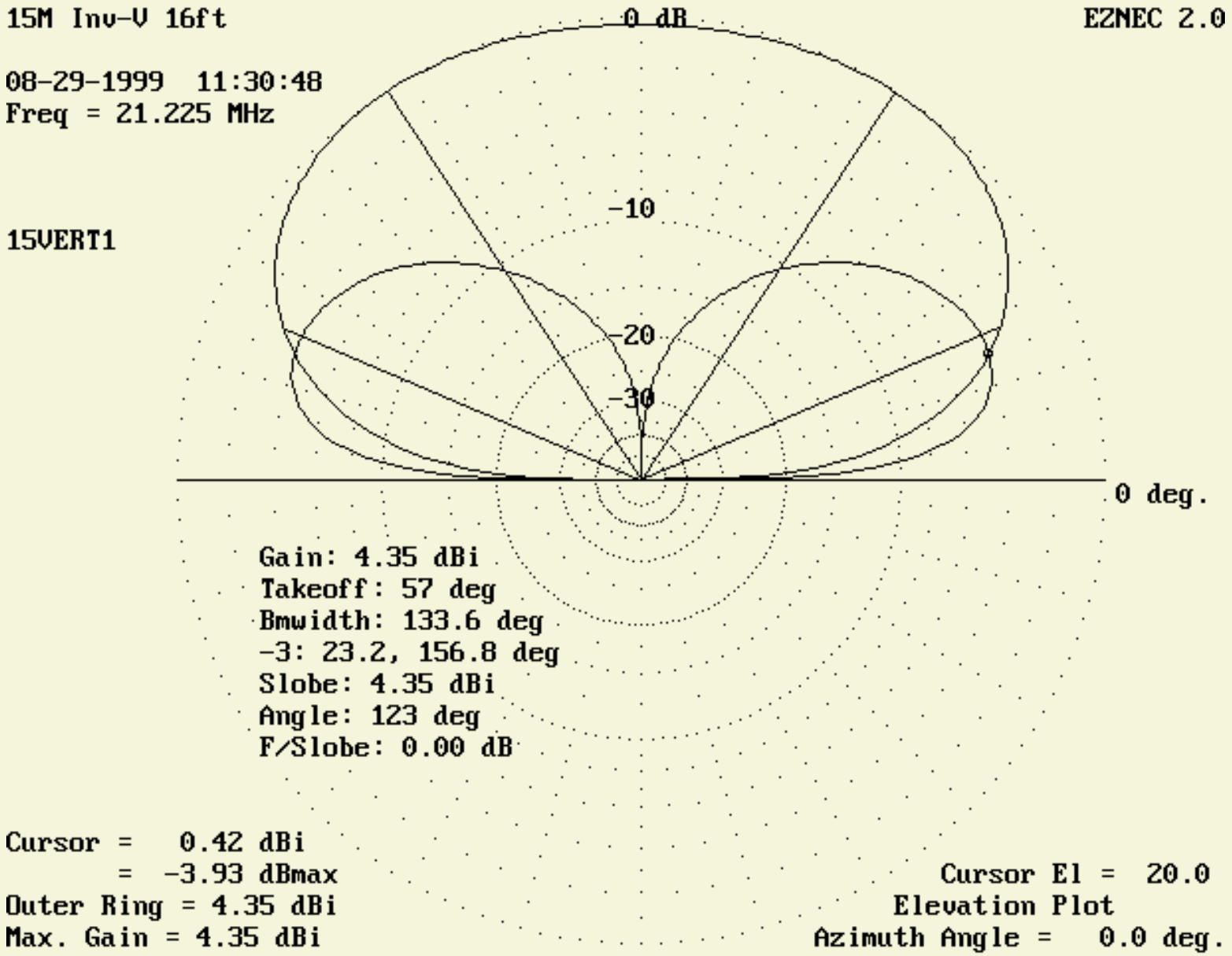


But notice the great similarity to the regular dipole. This is a cute antenna. It has one advantage. Since it assumes a slight droop in the upper wire, it supports the coax better than the regular dipole. The regular dipole was modeled with a droop due to coax or feed line weight. With an upward angle of 5 degrees on each side if center. This made a slight improvement in the pattern of the regular dipole.

So we decided to simplify things. With an inverted V you only need one support. Half the things to drag along. But there is a serious problem. As soon as the dipole started to droop a significant amount downwards at the ends, cancellation attacked. The sad truth is the inverted V is probably a bit worse than the vertical:



Let's compare the two on the same scale on the same plot:



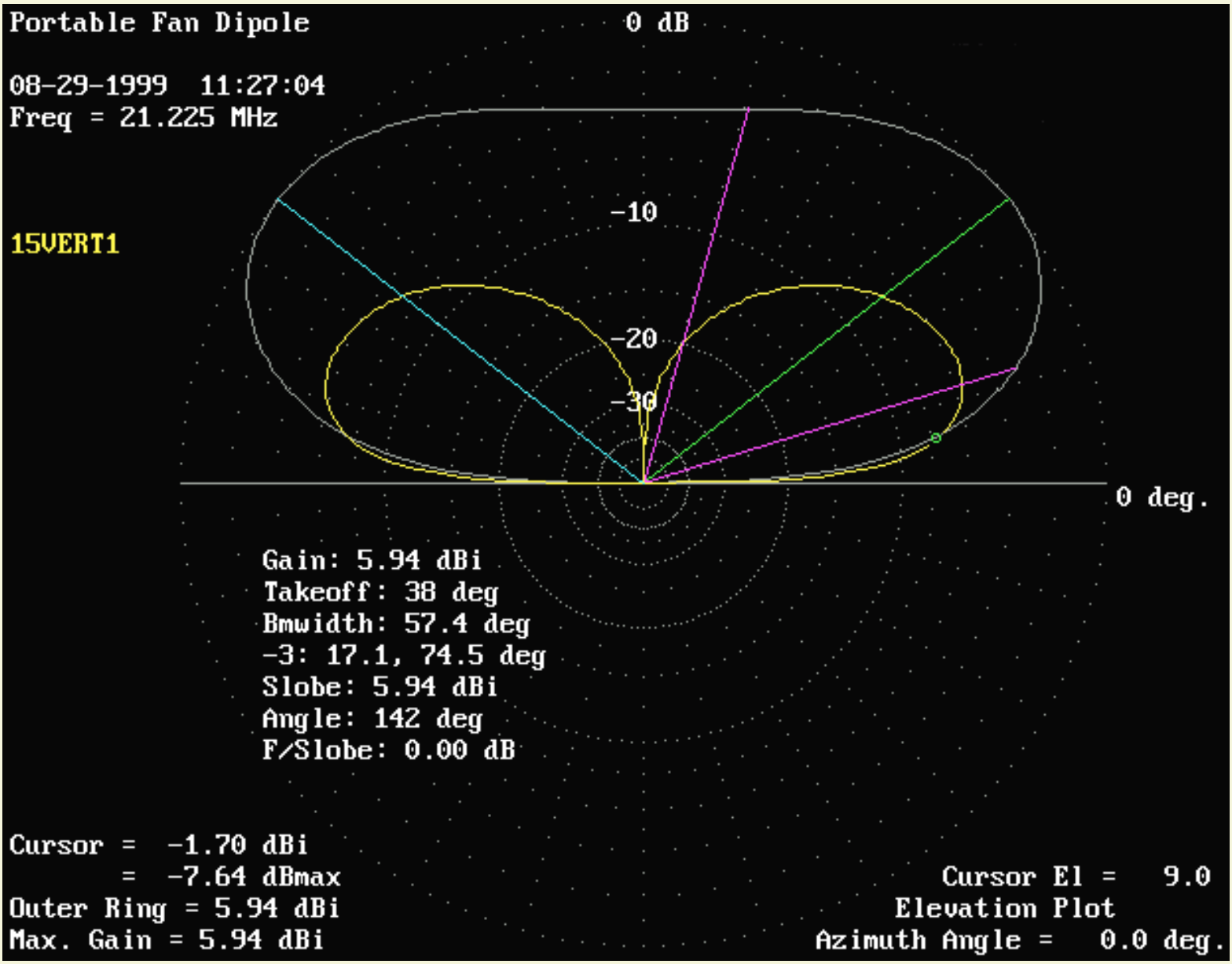
Note that below 20 degrees the vertical is better, enough better to call it significant perhaps. So, the sorry truth is that in spite of how pretty the vertical pattern is, and how nice it looks, and how primitive the old center fed dipole is with its simplicity, the dipole appears to be the best selection.

So why the great reputation of the vertical as a DX antenna? Probably for what it throws away. The pattern of the vertical will make it reject signals coming via NVI and high angle skip. This means it will reject signals from 50 miles to 400 or so miles out. If you live in California this means that lots of DX signals buried in local QRM might suddenly become readable as the vertical rejects the local stuff in favor of the low angle signals.

The vertical with ground laid radials is easy to set up and self supporting. When done this way it is a compact antenna to take along. The problem comes when you try to do better. You can use a pair of verticals and phase them. But then you have the same double hardware of the two poles supporting the dipole. The final sorry result is that phasing verticals is a bit tricky for best gain, you need lots of radials under both verticals to get them to work well and you might get about 3 dbi gain. The dipole is very easy to set up and tune and get the predicted performance out of. And it may well outperform your fancy two phased verticals unless you get them just right.

The support poles can be quite light, of nesting aluminum, relying on small guy wires. The dipole itself can be designed to be a light as possible as well to reduce strain on the poles. 16 feet should be quite manageable with careful design, making a bundle short enough to qualify as "fishing rods" a category recognized by airlines.

Finally, one last look, this time at the Fan Dipole vs. the optimized 1/4 wave vertical with 16 elevated radials. The cursor is placed at the intersection of the two figures and the lower left hand corner has the gain at that point:



In summary:

At one half wave above ground, a simple resonant dipole will significantly outperform the best installed 1/4 or 5/8 wave vertical.

At one quarter wave above ground, A simple resonant dipole will outperform a quarter wave vertical for NVI or local communication. The vertical if well installed over a decent ground structure, will outperform the dipole

for radiation angles less than 30 degrees.

Below one quarter wave above ground, the simple resonanat dipole will suffer significant loses, but still outperform a quarter wave vertical for NVI operation. Even a shortened loaded vertical, well installed, will outperform this low mounted dipole for radiation angles below 30 degrees.

An inverted V has both problems, the self cancelation of a vertical and the low mount pattern distortion of a dipole. But at 1/4 wave center mounting height it would be a good compromise between a quarter wave vertical and a horizontal resonant dipole. It would be much easier to make efficient if one is not judicious about the ground screen creation for the vertical.

Thus selection depends entirely on the height available and the number of supports. With two supports 1/2 wave tall, a resonant dipole, especially a fan dipole, will outperform the vertical in all cases.

With less than 1/2 wave of support/vertical pole, a carefully installed vertical will outperform the dipole for DX, but be much less useful for local NVI communications.

The inverted V includes the worst of everything, and is thus generally useful with a single 1/4 to 1/2 wave center support.

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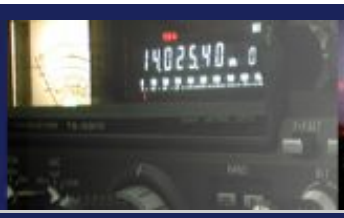

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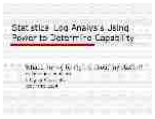
These include interesting tidbits written or uncovered by YCCC members.

First Annual DXpedition Award Program (MSWord) (PDF)
Award Policy (txt)
Award Application (txt)
by Jim McCobb, W1LLU

Log Analysis

80M Log Analysis (PPT)
by David Jordan, K1NQ

Statistical Log Analysis Using Power to Determine Capability (Rev 1) (PDF) NEW
by David Jordan, K1NQ



Antennas and Propagation

Compact wire Tri-band Yagi for 10 -15m (HTML)
by David Jordan, K1NQ (updated 22-Mar-2004)

10/15 M Dual Band Dual feed 26-foot Boom (PDF)
by David Jordan, K1NQ

What is VOACAP Trying to Tell Me? (PDF)
by Dean Straw, N6BV

Why We Stack 'Em - Covering All the Angles (PDF)

by Dean Straw, N6BV

Thrust Bearing FAQs (MSWord) - Hints and Kinks for your thrust bearings
Compiled by Fred Hopengarten, K1VR

Double-L For 80/160 (HTML)
by Don Toman, K2KQ

Two Wire Beverage (HTML)
by Jeff Parker, KA1GJ

Four-Square Antennas

Four Square Experiences (PDF) (HTML)
by Jack Schuster, W1WEF and Tom Frenaye, K1KI

Using a 4 square Vertical Phased Array to improve your 80 and 160 meter signal - without a Yagi! (HTML)
by Tom Frenaye, K1KI

FVR Spitfire - A Poor Man's 160 Meter 4-Square (HTML/PPT)
by Fred Hopengarten, K1VR and John Kaufman, W1FV
FVR Spitfire FAQs (MSWord)
by Fred Hopengarten, K1VR and John Kaufman, W1FV

Equipment

Improving the AL-1200 Relay Speed (HTML)
by Tony Brock-Fisher, K1KP

Operating and Tactics

The Thrill of it All (PDF) - Contesting isn't just for contesters (from QST) by Jack Schuster, W1WEF

Contesting Phrases (HTML) - Several contesting phrases are given in several languages
by Fred Hopengarten, K1VR

Sleep Deprivation Strategies (HTML) - minutes from 1988 presentation by Dr. Scott Johnson, NW1I - edited by Charlotte Richardson, KQ1F and Fred Hopengarten, K1VR

A Sleep Strategy for DX Contests - Randy Thompson, K5ZD

Dayton 1999 - Two Views

YCCC At Dayton (HTML) by Jeff Briggs, K1ZM

Dayton 1999 Recollections (HTML) by Jack Schuster, W1WEF

CQWW CW Recollections (HTML) by Jack Schuster, W1WEF

IARU 97 CW (HTML) by Jack Schuster, W1WEF

A Photo Tour of W1KM (HTML) by Tom Frenaye, K1KI

K1KI Station Profile and Pictures (HTML) by Tom Frenaye, K1KI

[Articles/footer_include.htm]

Antenna System Evaluator

This JavaScript program evaluates the performance of your antenna system, computing the consequences of SWR and feedline loss. 73 de Ron

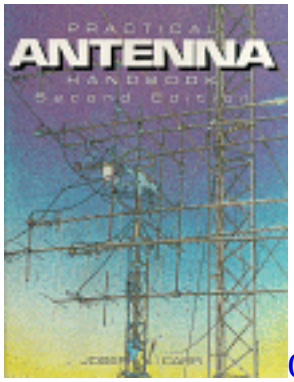
JavaScript PROGRAM by n6ach@comcast.net
original program concept done in CPM basic by KC6A and N6NB.

Translate this page automatically.

power from xmitter in watts	-1VSWR @ xmitter
feed line loss in db(must be in line with other values)	Antenna gain in dBi: Enter 2.2 for dipoles; add 2.2 for antennas rated in dBd

---

reverse power at xmitter	-1VSWR @ antenna
forward power in watts at antenna	reflected power in watts from the antenna
true power in watts into antenna	power in watts lost in feed line
db line loss including vswr effects	EIRP in watts



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If you have any questions, comments or you want to see something else added please e-mail me at n6ach@comcast.net

73 de Ron

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- Yagi-Uda 2M 12-L Beam
- The Biloop & No Capacitor
- Extend the 2-M OWA Family
- Grounded Horizontal Loop
- New Maxwell Equations Evidence
- *Read Stone's Throw! for News ~ Click Here!*
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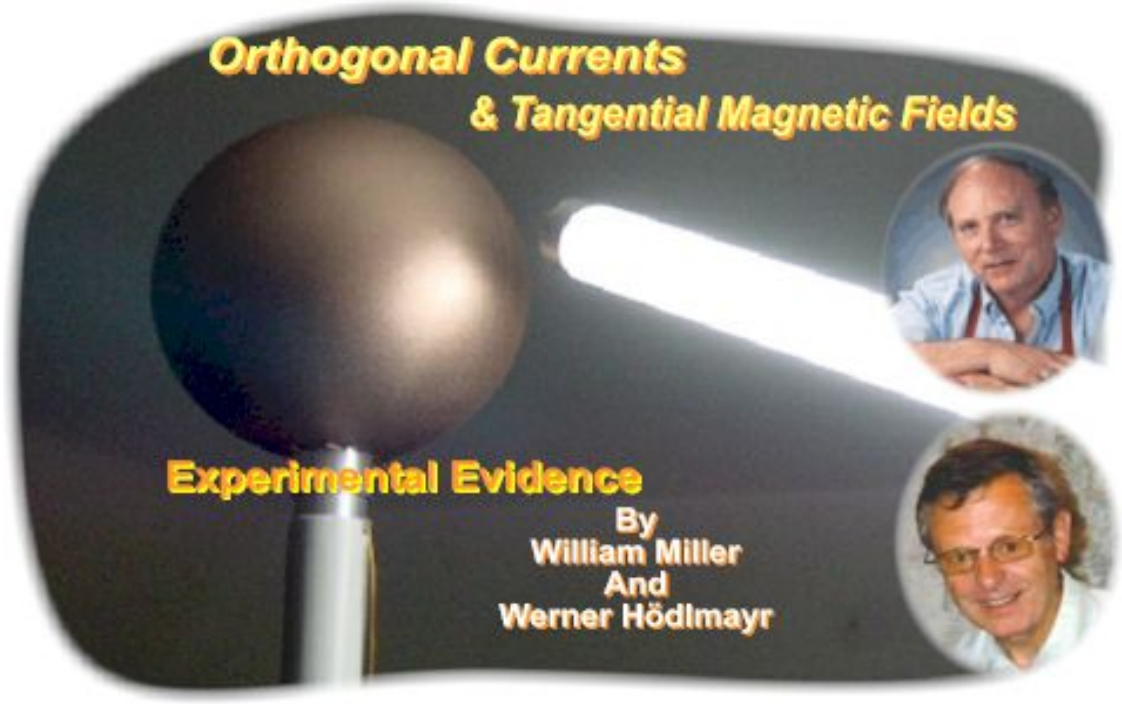

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Read Our Monthly Columns!

Antenna Modeling

This is a regular and popular monthly column by L.B. Cebik, W4RNL Because computerized antenna modeling has become widespread, and its popularity as a design tool continues to increase, this series is devoted to helping readers get the most from the design software used. The articles focus upon the use of NEC and MININEC, along with useful adjunct software as well.

From The Shack

This column is primarily for "Guest Editorials" to provide a podium for our readers to voice their opinions to the rest of the world too. This is a chance for readers to get on their "soapbox" and speak about antenna and radio-related subjects. Don't miss these interesting views about anything and everything about radio and antenna systems! Now, what have YOU to say??

Ham WorkShop

Ham WorkShop, is also another regular monthly column filled with a variety of "RADIO-STUFF" of value to almost everyone in amateur radio from Novice to Extra and those just beginning to take up this special hobby. This includes subjects, but not limited to: VHF, choosing the right antenna, coax cable, small to mid-scale construction projects in a practical manner, use of test equipment, etc. It is also meant to help readers become more familiar with the technical jargon *and* the fun side of radio.

Stone's Throw!

***Stone' Throw!* a monthly column by *antenneX* publisher, Jack L. Stone, among other things, is to keep the readers informed about our progress, new developments, plans for the future, and to introduce the authors and their subjects each month. Also, our main slogan around here is "we aim to please", so this serves as a place for the readers to tell the publisher what is wanted or at least make suggestions. Just remember, the publisher is only a *Stone's Throw* away! Go in for a visit and read this month's column.**

Propagation

***Propagation* another monthly column by Marcel H. De Canck, ON5AU of Belgium. Signal propagation is a subject that is one of the most basic ingredients of radio and is something everyone in radio should know about in order to maximize communications in the most effective way. It's not enough to have the best equipment and the best antenna if you are trying to send out a signal against a brick wall. Conversely, one may possess a very crude rig, running low power, but yet transmit/receive a signal to great distances with ease, simply by making use of a thorough knowledge about how signal propagation works within the environment. Follow this column and learn more about propagation!**

FEATURE ARTICLES THIS MONTH IN THE LIBRARY

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Extending the 2-Meter OWA Family
Part 3: Increasing OWA Gain
vs. Preserving Sidelobe Suppression

By L.B. Cebik, W4RNL

In our preceding discussion, we developed tentative answers to two out of our three inquires that emerged from the extension of the OWA 2-meter family to 20 elements. By way of quick review, the low gain but high sidelobe suppression of the 20-element version of the array left us with these questions.

- **1. Which of the two OWA design concepts--the core or the method of adding new elements--has the dominant effect on side-lobe development?**
- **2. Is there a method of obtaining more gain from the OWA series without sacrificing sidelobe suppression and attenuation?**
- **3. What role does element population play in sidelobe attenuation?**

Building a CCD Antenna

By Joel C. Hungerford, KB1EGI

This month, I continue looking at a thing I recently heard about on the antenna discussion list — a so-called CCD antenna. CCD stands for "Controlled Current Distribution." It consists of a string of series tuned circuits and short bits of wire between them. The circuits are all tuned to the lowest frequency to be used. The current at each circuit can be set by the reactance of the inductor and capacitor. It seems a natural for the combination of the interlaced wire capacitor I used to tune the loop, and small coils, all wound on a long PVC pipe. What if one used a modest number of tuned sections, say 5 to 10, and increased the reactance as one progressed from the feed to the end of the pipe, following an approximate tangent curve. Could this simulate the current in a vertical, but in a shorter distance? Would it radiate? Just how does the tuning behave if I string together several sections all tuned to the same frequency, but with a different L/C ratio in each section?

So, this month I set out to build a CCD antenna, consisting of several series tuned circuits all resonant at the same frequency, but with each section L/C ratio varying along the antenna to mimic the effective impedance of a dipole: high impedance at the open end, and low impedance at the center.

Yagi-Uda 2-Meter 12-Element Beam Design

By Fred M. Griffee, N4FG (EE Retired)

In this article, a 2-meter 12-element Yagi antenna will be designed using optimization in lieu of the Optimized Wideband Antenna (OWA) approach (Ref 1, 2, 3). The optimization approach I shall use follows the conjugate-gradient optimizer type found in many programs such as YO by K6STI. However, the more interesting approach may be the generic optimization approach, but to date I find it to be overly lengthy, cryptic, very time consuming, and difficult to implement. I find the conjugate-gradient approach creates designs through its progression that follows user defined gain, SWR, impedance, F/R ratio weighting, and element characteristics (diameter, wall thickness, material, etc..). Any of the mentioned characteristics can be altered where the azimuth coverage is assigned with respect to 180 degrees (N – 180 degrees). The DL6WU-gg program, for example, creates side lobes that start at 45 and 315 degree E-plane azimuth so the range will be assigned 45-180 degrees.

THE BILOOP – A LOOP TUNABLE WITHOUT THE USE OF CAPACITORS

By Claudio Re, I1RFQ

Recently, on the *antenneX* antenna discussion list, several contributors expressed interest in learning more about improving the performance of compact loops. This was prompted, in part, by the introduction of some new antenna concepts based on recent articles on the "Cubes Family."

The most common question seemed to be, "What techniques are available to significantly improve the performance of compact loops?" Over the last few years, the author has spent quite a lot of time reviewing this subject from the point of view of the amateur radio operator. This article contains some of the results of these studies.

THE GROUNDED HORIZONTAL LOOP
A single antenna from 10 to 160m
Or, the TTT: Top-Band Top-Fed Top-Cap Antenna
J.M. Bourdereau, MD, F5LCI of France

Author J.M. Bourdereau says the "horizontal loop antenna" has many advantages, when there is enough area. Grounding at a voltage node does not affect its characteristics, allows static drain, and allows its use as a top-loaded top-fed vertical for lower frequencies. This idea seems confirmed by computer modeling. Thus, we have (at least) two antennas in one.

ORTHOGONAL CURRENTS
&
TANGENTIAL MAGNETIC FIELDS
Experimental Evidence
By William Miller & Werner Hödlmayr

In the 1860's, James Clark Maxwell postulated the existence of Displacement Current as the mechanism whereby Alternating Current (AC) flows through a capacitor. It forms the keystone in a series of equations — often called Maxwell's Equations — that are the basis of all Electromagnetic theory.

About a year ago, *antenneX* published the author's 2-part article called, "Displacement Current Does Not Exist." The article was prompted by the apparent failures to perform as claimed found in two classes of antennas — the CFA (Cross Field Antenna) and the EH. The inventors of both antenna types claim that they "use" displacement current as an integral part of the radiation process.

July's issue of antenneX (July 2004) featured an article by William Miller entitled, "How A Capacitor Really Works." In this article, Mr. Miller postulated an alternate explanation for Maxwell's "Displacement Current." Displacement Current is the key ingredient in the equation set that relates the total Magnetic Field from an AC conductor to the magnetic field generated by current flow and electric field.

In his pursuit of working proof, Bill has recruited the very able assistance of Werner Hödlmayr, well known for his previous work on the MicroVert, TeslaVert, Fractals and others. The latest of Werner's TeslaVert versions was chosen for some experiments while applying Bill's theories. This month's joint article discusses Bill and Werner's initial experiments and the early evidence realized from them.

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- NEC4WIN95 Software Review
- Economical Vertical Antennas
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- Analysis of CFA Field Strengths
- Microwave Absorption in Humans
- CFA & DLA Discoveries Using *PSpice*
- Mobile Operation in Outback Australia
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- Effects of VSWR on Transmitted Power

Professional RF Network Designer: A Short Review

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? need a program for antenna design and antenna plotting
? don't know what antenna is best for hamsats or others
? have a busy mobile signal on all bands
? need an inexpensive directional antenna for 10 meters
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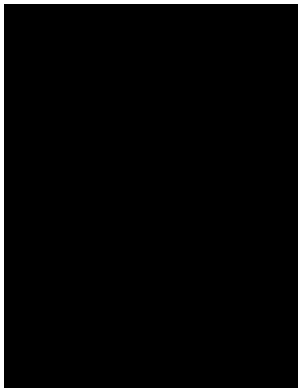

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Last modified: July 04, 2004

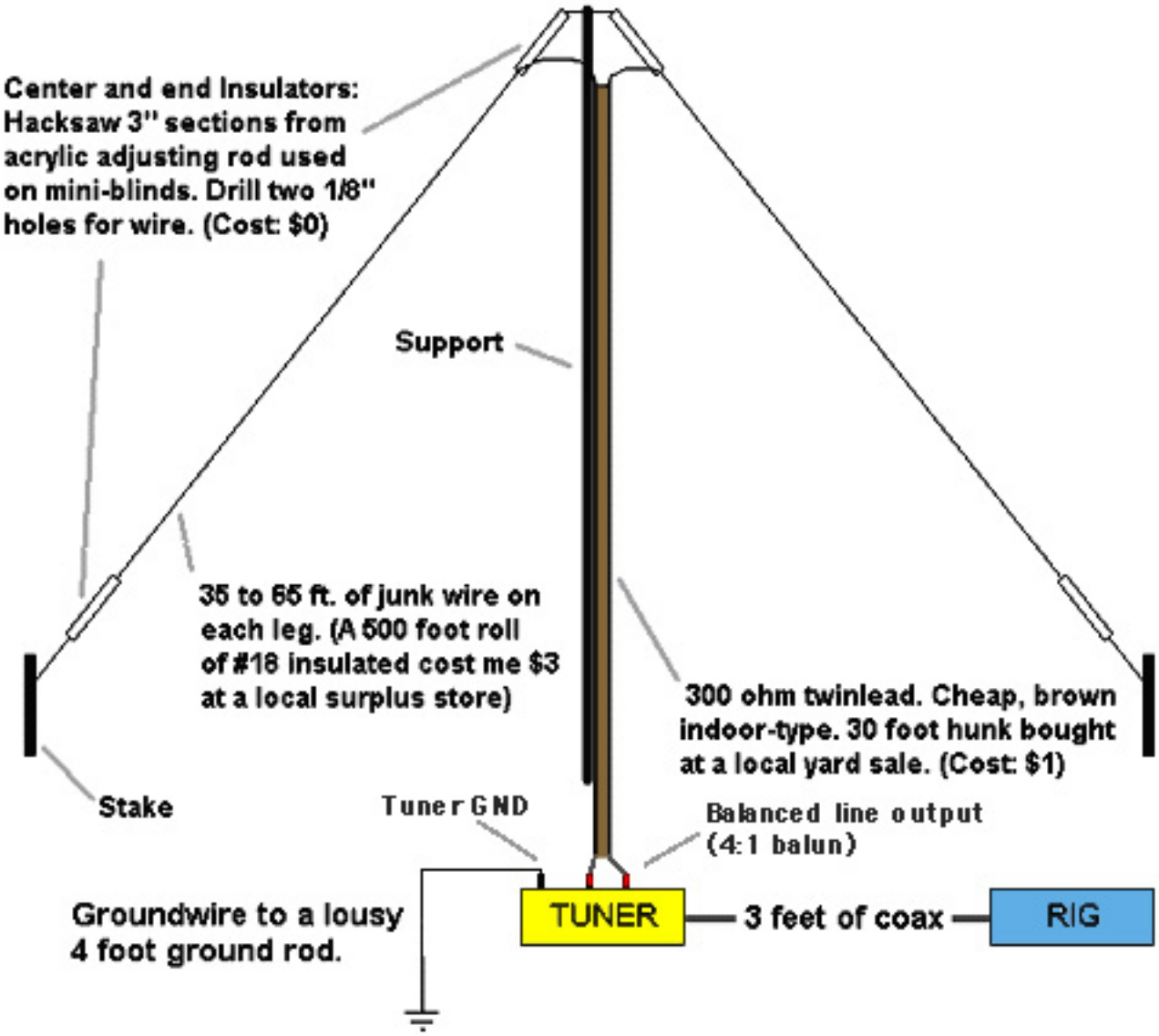
THE $4 SPECIAL

by Joe Tyburczy, W1GFH

Sure, you can find "all-band wire antennas" for sale in the back pages of Ham magazines costing $150 or more. But beware: *Marconi spins in his grave everytime a ham buys an aerial instead of building it!* Look, the plain and simple truth is that wire antennas for the HF bands were intended to be *hand-made* and not store-bought. Untold generations of intrepid Radio Hams have fashioned their own equipment out of spit and baling wire. In this world of microprocessor controlled micro-rigs, this may be your only chance to build something and actually see it work on the air. Think about it.

Another bonus of "rolling your own" antenna is that it costs you next to nothing. Don't be intimidated by SWR, either. Your rig will not blow up and kill you. Most modern rigs will politely refuse to transmit into a really bad match. A perfect 1:1 SWR is for sissies, anyway. All *real* hams have conducted perfectly good QSO's at 3:1, and even 10:1 at some time or another. Anyhow, I recommend a tuner. This, you can buy over the counter with a clear conscience.

I am a big believer in balanced line vs. coax. The basic "W1GFH $4 SPECIAL" shown below is a variation of the type of exceptional skyhook I've been using for years.



I've tried the commercial 450-ohm ladder line, but prefer 300-ohm TV twinlead, and the cheaper the better.Forget all that crap about impedance, wavelength, and velocity factor. What you really need to concentrate on is getting an interesting set of *antenna insulators*.

Back during the disco era when I first got on the air, I got a pair of really cool antique *pyrex antenna insulators* from a flea market table in Derry, NH for 25 cents each. They looked like the kind Hiram Percy Maxim used in 1910, and seemed able to pull in exotic DX all by themselves. The other day I

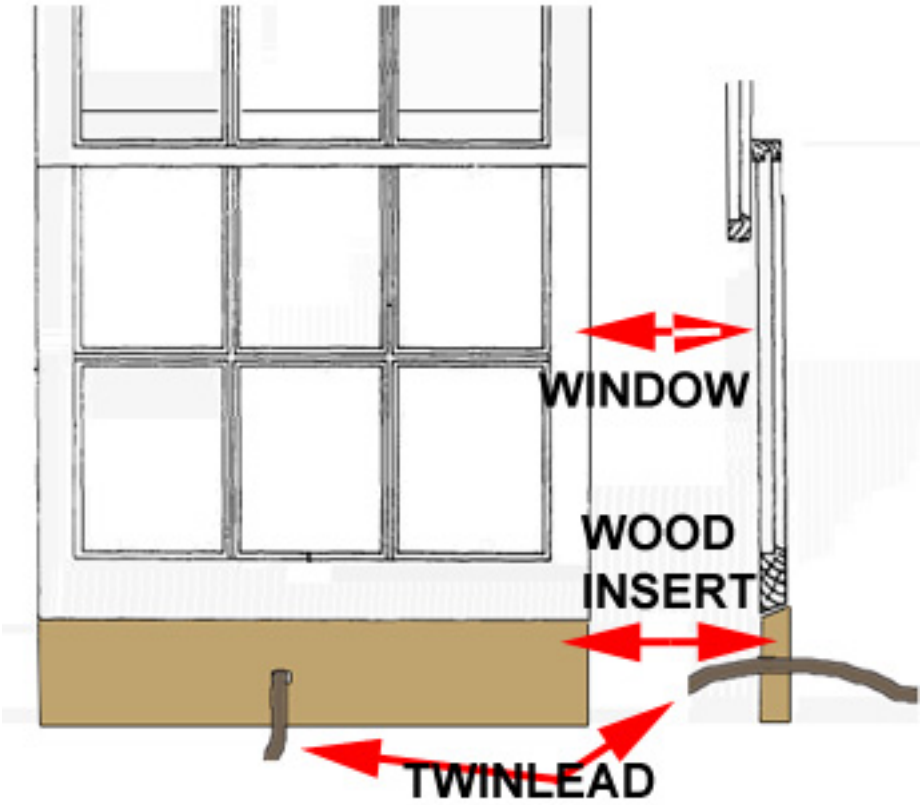
found out that Radio Shack wants $5 apiece for insulators made from some kind of white plastic crap. So I improvised my own by sawing up pieces of an acrylic adjusting rod from a discarded miniblind. I think Hiram would've been proud of me.

The tuner was an MFJ-949C I got at a swap meet for $15 a long time ago and has connections for balanced lines on the back. A link-coupled balanced tuner arrangement like the Johnson Matchbox would be better, but use what you have. The idea is to *get on the air with a decent signal for cheap*, just like they did in the early days of ham radio!

I had a 65ft. per leg version of this antenna working in Massachusetts, and it'd tune up on all bands 80-10. At my Burbank, California QTH, I used a 35 ft. per leg version, and it tuned up on 40-10. By the way, you'll notice it's an inverted vee --- a real advantage if you don't have room for a full-on dipole in your yard. If you still don't have room, bend and angle the legs to fit the space you've got. Antennas gently bent into Z-shapes still work fine!

THE *MAGIC* OF TWINLEAD

The feedline comes straight in thru a window sash. The nice thing about the 300 ohm twinlead is that the small stuff only needs about 2" separation from metal objects in its path. (OK, so TV Twinlead won't take a kilowatt, but the 450 ohm stuff you can obtain at ham stores will. You just need to be more careful with routing it)



You can get a 4:1 balun to make the transition from the balanced line to a length of coax, so you can bring the coax into the house via a properly sealed-up feedthough hole to avoid winter drafts if you'd like. If you really want to get nutty, try using 110VAC lamp cord ("zip" cord) as a feedline. Yeah, it'll work as a crude balanced line, believe it or not. Impedance varies, but is usually "close enough" to work. *Improvise. Experiment.* Take notes of what works and what doesn't. This is what ham radio is all about.

Many of you will recognize this antenna as the venerable "double zepp" aerial, a variation of the "end-fed Zepp" -- the skyhook responsible for the dramatic Hindenberg tragedy in Lakehurst, NJ. It seems the blimp's radio op decided to work a little DX while waiting for landing clearance. He sent out a few CQ's. Unknown to him, the ladder line had twisted in the breeze, shorting the bare conductors. A brilliant spark flared

up, and....well, that's another story altogether.

To see an "end-fed Zepp" version of the $4 Special, click here.

Alas, I never had the height to make this antenna perform the way it should. The one in Mass. was up 50 ft. and worked terrific DX. The one I have now is only up 30 ft. and gets average results. It won't outdo a yagi at 100 feet, but what will?

And for $4....who can complain?!

ADDENDUM: November, 2002

Since writing this article, I've gotten a lot of questions. Many are curious about the end-fed Zepp. I suggest you go to L.Cebik's fine page on this subject for an explanation of the practicalities of such an antenna:

http://www.cebik.com/gup/gup12.html

The other question deals with feedline lengths. Is there any 'ideal' length? Yes and no. Some feedline lengths will

present an extremely high impedance to the tuner on certain bands. Each installation is different, but here are some rough guidelines that may help:

Start by trying a feedline listed in the lengths below. It may take some trimming or adding of feedline to work well on the range of bands you want to cover. The worst possible feedline lengths are shown in brackets:

If Ant is 120 ft per leg it will cover 160 thru 10 meters. Feedline of 40-70 or 150- 190 feet suggested. [Avoid lines around 120 or 240 ft]

If Ant is 65 ft per leg it will cover 80 thru 10 meters. Feedline of 25-40, 80-100 or 140-160 feet suggested. [Avoid lines around 60, 120, or 180 ft]

If Ant is 33 ft per leg it will cover 40 thru 10 meters. Feedline of 40-50, 70-80, 100-110 or 130-140 feet suggested. [Avoid 30, 60, 90, 120 ft]

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Messages: **56**

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	Jan	Feb	Mar	Apr	May	Jun	Jul	Aug	Sep	Oct	Nov	Dec
2004	380	424	485	392	444	439	409	392	50			
2003				501	647	486	350	263	234	280	375	445

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Members: **4625**
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Language: **English**

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Welcome to G4VGO's Kite and Balloon Antenna Site

Featuring Limited Space Solutions for the Low Bands

Some of the Top Band Antenna Information You Will Find Here:

Page Two, My Kite and Balloon Lifted Antennas
Page Three, All About 'Aerial' Antennas
Page Four, Kites as Lifters, The Background
Page Five, Balloon Lifted Antennas and General Flight Safety Issues
Page Six, Winches and Rigging for Kite and Balloon Antennas
Page Seven, The Full Wave 160 Delta Loop...It Worked Well.
Page Eight, Miscellaneous Aerial Antenna Information

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For now, the QTH is Singapore.

The call is 9V1GO.

My name is Bob, and I have been an amateur radio operator since 1962. My activity on Top Band goes back 20 years.

Now, I am living and working in Singapore. This past season (2003-2004) 91 DXCC entities and 30 zones were worked on Top Band. Now there are 92 DXCC countries in the 9V log,

thanks 4X4DK. Conditions have been getting better the past few weeks, and it is nice to know that the new season is not too far away.

If you think you have worked me from **9V1GO**, you can check the log at **http://dx.qsl.net/logs/index.html** and look under my call for the logs from here.

The 160 metre transmit antenna is a half sloper at 40 metres. I have one receive antenna that has out performed pennants and flags. It is a full sized half wave dipole two metres off of the roof of the 13 floor apartment building. I also have a tuned co-ax loop that occasionally out performs both the low RX dipole and the TX sloper, depending on conditions. Watch for Top Band activity from Singapore on 1814.5 and 1822.0 at my sunrise and sunset. In addition, CW on 80, 40, and 30 metres will get a lot of attention as the S9 QRN on 160m in Singapore is hard. Up to more than 100 DXCC on all the rest of the bands but 10 and 12 metres. Email me for a sked if you need 9V on any of the low bands. **Please QSL DIRECT ONLY to my manager OK1DOT -** Petr. His address is on the QRZ.COM web page.

Now, for a little about G4VGO at home:

In "G" land, Top Band is my O-N-L-Y band, and I have been active in Europe since 1984 on 160 meters. Shortly after coming to the UK in 1980 I was issued the reciprocal call G5EPD. A few years later it was changed to **G4VGO**. That has been my call in the UK since then. In 1985, I managed to confirm 102 countries and received 160 Meter DXCC Certificate #269. Since then the total has increased to 224 with 39 zones. The entire country total has been on CW.

I am now fortunate enough to live on a piece of land with wide-open space (when I get home). The farm is in the middle of countryside, not far from the North Sea coast and with more than enough room for beverages, and tall masts. Although the necessity for kite and balloon lifted antennas is not so acute with space in every direction and no neighbours to worry about, I still really enjoy putting the wire high up in the air. The chances of increasing your score in a Top Band contest or pushing the country total higher go up by about the square of the height of the antenna on 160.

My job is a telecommunications consultant and I work mainly with mobile networks. This profession has taken me all over the world. I have deployed networks on all continents but Antarctica and came close to there when we put a site in very southern Argentina. It is fortunate that my hobby and my work both involve radio and I enjoy both very much.

In 1997, I was **EI7IU** in Naas in county Kildare, just outside of Dublin. In seven months I managed to work more than a hundred countries on Top Band including Myanmar and a few other rare ones with 150 watts and a 14 metre 'T' antenna.

For about eight months in 1999 I was **LU/KY0C** in Buenos Aires while working there. Conditions were terrible in the noisy city environment, but regular QSO's with W8JI and other USA big guns were the norm. It was good to get back to the UK, to low noise and better propagation.

In 2001 I spent five months in Israel as **4X/G4VGO** and with 25 watts to a full sized 160 metre sloper and two Pennant receive antennas managed to work more than fifty countries including the USA and Canada. Again, Israel was a very noisy location with poor summertime propagation.

I left **SM Land (SM0/G4VGO)** after about six months there in the last half of 2002. While staying in Kista near Stockholm I worked 42 countries with a marginal antenna and very low power. I didn't have such good antenna possibilities, but as the nights got longer, DX picked up a bit.

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The Home Shack

This is the home QTH operating position. I have tried every top end transceiver in the world, and for several years used FT1000Ds as the mainstay. For a while, I used an FT1000MP, but went back to the 1000D. I also bought an ICOM IC746Pro to try, and after exhaustive trials on Top Band, sold one of my FT1000D's and opted for the 746Pro. It was a good transceiver, but then I had a chance to try the Ten Tec Omni VI. After a few nights on 160m CW, I decided I should have gone with the Omni VI years ago....it is AWESOME on CW receive. For CW, I use the Omni VI, but for SSB, the D104 and the TS830S are a very good combination. The old Kenwood has done well in many SSB contests on Top Band.

My latest purchase here in Singapore was a Elecraft K2/100. With the latest firmware and keying mod it is rapidly becoming a contender for my main radio. (Although the Icom 756 Pro II I am also using does have a pretty good receiver on the low bands. However there are no

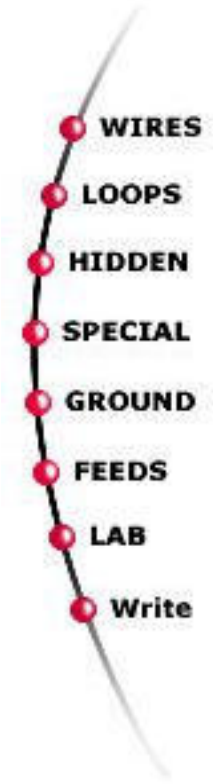
strong signals nearby to contend with at my 9V QTH) European Top Band activity is a challenge for any receiver, so the K2 may be the new rig of choice upon the return to G4VGO.

The Dreaded Key Clicks

Most stations I hear on CW have key clicks. After participating in many contests on CW and suffering the clicks, I decided to get rid of all of my Yaesu radios and change to the Omni VI and the trusty old Kenwood TS830S. The TS830S has **NO clicks** or phase noise**,** and the RF speech processing in the 830 makes it one of the best rigs for that mode ever made. The Omni VI has had the key click mod, so it is a lot better than anything I have used in the recent past. All of the transceivers I am using now do NOT cause the obnoxious and very broad key clicks that the Yaesu FT1000 series (**ALL of the FT1000 series**) cause. In a pile up, the un-modified Yaesu rigs sound like a convention of castanets clicking up and down the band several tens of KHz. Really rotten keying. Add several tens of dB of phase noise and you have a proper mess on Top Band.

There is an excellent "must read" article by Tom W8JI on CW Key Click's (Spurious Emissions) caused by Yaesu Top-of-The-Line FT-1000 Series HF Rigs. This includes, the FT-1000MP, the Mk 5, the MK 5 Field and FT-1000D Models. The receivers may be mostly OK, but the CW transmitters are ***CRAP***!!!!!! I see more and more of the FT1000 variants for sale used now so maybe someone is listening. Or, maybe they are tired of hearing complaints about their terrible keying from those they clobber with the clicks. Anyway, it is clear that Yaesu has "zero" concern about good engineering practices, customer service, quality control and spectrum management. We should all think twice before buying one. Can Yaesu say "Lost Revenue Stream"? The new Icom radios have variable keying rise times, as does the new Ten Tec Orion, why can't Yaesu make a good CW transmitter?

I use Beverage and Pennant antennas for receive. An east Beverage on Asia and the Pacific one and a half wavelengths long on 160.... a pair of phased end-fire Beverages to the northwest on North America, one and a half wavelengths long and a south Beverage one wavelength long take care of the main DX directions. The Pennant receive antennas look southeast and northeast as 'gap fillers' on Africa and deep Asia. Beverages and Pennants at this quiet location have given me a whole new level of weak signal capability that I had forgotten existed. They also give me a front to back ratio that gets rid of the European key clicks that can drive you nuts.



Balloon or kite antenna? Why not

- First thick nylon line, then copper wire
- Be extremely careful about static
- My balloon and kite antenna system
- Kite antenna - watch out!
- Balloon antennas - a theoretical approach
- Excellent supports for LF antennas

First thick nylon line, then copper wire

From: Hugo Caron (hugo@infobahnos.com)
Date: Nov 4, 1996
Original source: Usenet's rec.radio.shortwave

Last month there was a discussion about balloon antennas, mainly about using 2 balloons instead of one for a vertical antenna.
Two persons mentioned their concerns after having read articles on the subject relating about wind-induced static shock hazard and reception high noise level.
One participant replied that if one would shunt the antenna with carbon resistor (~4.7k-22k) to local ground, high enough not to upset antenna caracteristics but still, low enough to discharge the beginning of charge buildup, one would be able to cope with this problem.

On the same subject, a second participant with actual balloon and kite antenna experience (300 ft - 100 meter) did report the reality of shock hasard, but eventualy got around the problem by shunting an antenna tuner inductor to ground or shunting the antenna directly with a 100k resistor. He reported that in these conditions, he was able to get signals that sounded great. Also in light wind conditions, he said that static was not much of a problem.

On the structural aspect, for security reasons, personally I would't try to tie a thin copper wire directly to the balloon, as some may have suggested. I'd rather use high strenght nylon or alike line (fishing line type for exp., I've seen some with 40 lbs capacity) to cary the all strain the holding line may be subject to in presence of higher ds than expect winds. The thin wire copper line would only be ' brought along ' for its antenna role. To avoid any wobbling of the 2 lines, I would knot, using short lenght of holding wire (trimmed after knotting), every 5 feet or so. For example, if one want to jack a 300 feet vertical up and chooses a

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nylon line of 20 lb capacity (lets say it's 2000 feet/lb (.0005 lb/foot) and AWG #28 copper wire at 2081 feet/lb (again, ~.0005 lb/foot) and 65 ohms/1000 feet (ARRL handbook), one would end up with a line at .001 lb/foot for a total payload of .3 lb. So the balloon would need, at least, this 'net' boyancy (preferably a bit more let say .4 lb). By net boyancy I mean the payload the balloon can handle, after lifting its own weight. A 3.5 to 4 feet diameter ballon should be able to handle this charge, depending on its enveloppe weight.

The balloon envelope material quality his also something one should consider before attempting any lift off. If not, one may experience a short lived project.
We've all air blew party rubber balloons to their max, just to realize that the next day they had shrinked to half their size. The reason being that, due to pressure differential between inside and outside of the streched enveloppe, air will just sift through it. Rubber is an elastomeric material build around long organic molecular chain attatched one to the other through ramifications (something like a tree), and there are 'holes' between these chain elements. Air (that is, molecular oxygen, nitrogen (and carbon dioxide if you blew the balloons yourself... pfiew!!)) eventualy finds its way through. So if molecular elements like O2 and N2, that are many time the size of atomic helium (He) can do that, you can imagine how helium would act considering the wide open barn doors these 'holes' are, relative to its size. I dont think one would be able to retain it for more than a couple of hours. So not any off the shelf stock will suffice.

I've done a few phone calls, a couple of weeks ago to see what's available, mentioning I was looking for a balloon to be inflated with helium, that was able to retain its content for at least a couple of days. The only source I found was the Party Decoration retailers. I eventualy found someone that told me they had 42 inches size balloons, the enveloppe of wich is made of unstreachable material (vynil???) made for that purpose (helium inflating) for $5. They would inflate them for $5 each. One can also get pressured helium portable refill thanks, good for ~25(?) of these balloons, at $60. They were not able to specify the enveloppe weight or the ' net boyancy '.
Another retailer told me the he could supply me with a 7 feet balloon (that's 9 times the volume of 42" diam.), again the unstreachable kind used for outside advertizing, hence relativaly long lasting, with a 4 lb lifting capacity for CAN$342 + helium...
Ooops, this time *my* budget was blown!

Be extremely careful about static

From: Jake Brodsky frussle@erols.com
Date: Nov 6, 1996
Original source: Usenet's rec.radio.shortwave

First: Hugh Caron's post concerning static charges on the balloon wire are accurate. I was one of the guys he spoke about and, yes, when flying kites with a wire antenna, one must be extremely careful to bleed off the static charge. My friend Tony, was practically thrown across the deck of his boat when he touched the ungrounded wire on one of our kite- antenna experiments.

Two: Helium is much easier to buy and transport than hydrogen. It may not be cheaper, but at least you don't get anything close to the HAZMAT paperwork you'd get if it were hydrogen.

Three: Although latex balloons do deflate more readily than, say, mylar balloons, the surface area to size ratio of an eight foot weather balloon will guarantee that it will stay up for quite a while.

Four: The FAA's relevant regulation on this subject of tethered kites and balloons is Part 105. When I last read it three years ago, it said that anything with a total weight of under five pounds empty is exempt from most regulations.

You do have to know where the airports are, and don't fly your kite or balloon where it may get in the way. A call to the local Flight Standards District Office (in the United States) will go a long way toward clarifying whether the place and altitudes you wish to fly the antenna are appropriate. Even if you might fly it so high that it could interfere, they'd be more than willing to accomodate you by issuing a NOTAM, or NOtice To AirMen, with adequate notice, and then issue you a permit for a certain time and place.

My balloon and kite antenna system

From: Jake Brodsky frussle@erols.com
Date: Nov 7, 1996
Original source: Usenet's rec.radio.shortwave

Since the discussion is going this way I'll go in to more detail as to how I handled the Balloon and Kite antenna systems.
First, all flights were made from a boat, a few miles south of the Middle River in the Chesapeake Bay. All lines were 400' or less.
The kites and balloons all had empty weights of under five pounds (thus not requiring notification under part 105). Prior to conducting these tests, I contacted the Baltimore Flight Standards District Office for advice on how to proceed.

We thought briefly about using a strobe light. The problem was two fold: first, the strobe itself sometimes put us over the part 105 weight limit (depending on which kite we used). Second, the strobe light could easily have been mistaken for a life-vest strobe light on the water

(because that's what it was). We decided to keep the line to about 200' instead, and to forget the strobe light.

There are few standards for this sort of activity. Kites and tethered balloons are usually not a problem for most aircraft, especially those over open water. I caution you, however, to know the area where the airports are, to know where the instrument approaches are and to avoid those places. This is something I'm already aware of, since I am an instrument rated pilot.

In any case, low flying aircraft don't often go over open water, so the risk of collision is low (the big sky, small target theory).
Obviously, these details are things you should consider well before purchasing that kite or balloon...

Kite antenna - watch out!

From: Glen Leinweber leinwebe@mcmail.cis.mcmaster.ca
Date: Dec 12, 1994
Original source: Usenet's rec.radio.shortwave

Using a kite, or balloon, to launch a long-wire antenna is a risky business, even on a clear day with no clouds around. The danger arises from the electric field between the earth and ionosphere. This field is about one hundred volts per meter (in the vertical direction), and is always present. Any antenna launched into the clear blue yonder will aquire a charge large enough to wipe out a final transistor, or a receiver front end.

So what's the difference between ordinary outdoor antennas and one attached to a kite?
Earth's electric field is easily distorted by objects attached to ground, like a tree, or mast, or house (your tower doesn't have a few kilovolts between top and bottom). A kite-bourne antenna protrudes into wide open spaces, where it very likely gets charged up. Your earth-bound antenna is attached to a mast or tree or house, where earth's electric field is greatly reduced.

Here are some precautions for kite experimenters, or anyone who uses temporary outdoor antennas:

Add a choke coil to the antenna connector between centre pin and ground. This will leak away any charge that tries to accumulate. I see that most modern rigs don't provide a D.C. discharge path from antenna connector to ground - this is dangerous to your equipment.

This choke won't save you if you connect a charged antenna to your rig after you've put it up: either discharge the antenna to ground or connect

the rig to antenna **before** putting it up.

A few years ago I put up a two-meter 1/4 wave antenna on top of my tower. The coax snaked thru the window had no connector. When it started snowing, I noticed a snapping sound every twenty seconds or so. The end of the coax was arcing over from the charge accumulating in the coax capacitance. That cable was being charged five or ten kilovolts in only twenty seconds!

Balloon antennas - a theoretical approach

From: Hugo Caron (hugo@infobahnos.com)
Date: Oct 27, 1996
Original source: Usenet's rec.radio.shortwave

The discussion about the possibility of erecting a balloon supported long vertical and/or inverted L type thin wire antennas has raised concerns about potential shock hasards and high noise levels resulting from wind-induced static. Solution replies were proposed, through the use of bleading resistor (and/or inductor) to a good ground, high enouph to discharge the static buildup at adequate rate while preserving electrical caracteristics of the antenna. r > 4700 ohms. inductor = ?

Concern was also raised about the trouble of putting up a 2 balloon setup over a single one as high (or even higher) vertical, on the base that inverted L would behave as top loaded vertical, which is omnidirectional by nature. On this, I replied it *may* and *would* only be worth the trouble *if*, considering the multiple wavelenght in hight and lenght for a large part of the SW band, such a setup would show *directivity* caracteristics that would approach that of a long wire. As I could't answer this question myself, I left it open for comment. If there were any, I have not seen them, yet.

Lastly, concerns were also expressed about the physical (structural) ' reality ' and security of such a setup. To this I brought the idea it should take the shape of a trapezoidal figure (as seen from the side). I also stated that I will try to determine through mathematical (mecanical) equations, proper sizing of such a setup along with constrains the wiring would be submit under different wind loads.
This is what will now follow.

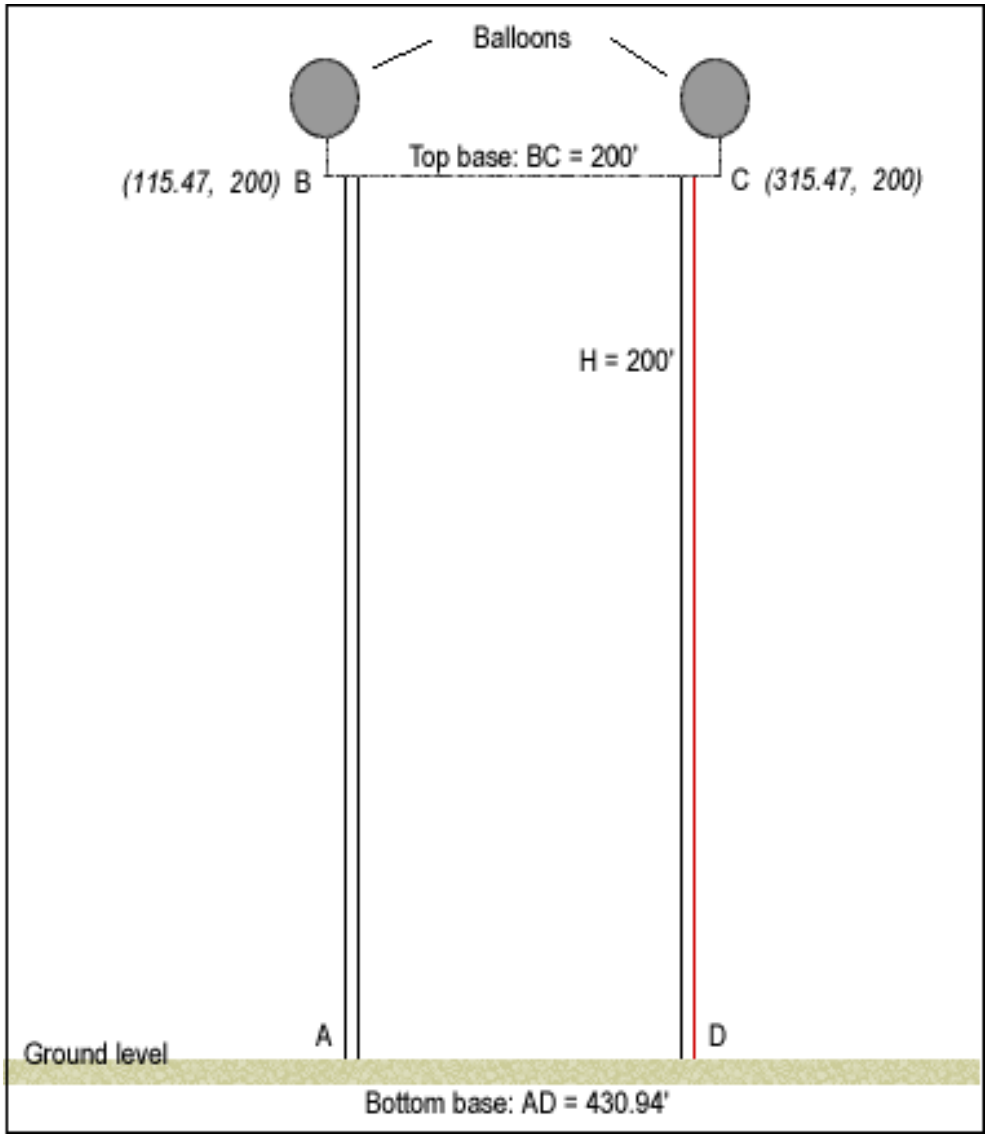
Sizing for still-air conditions

Material specification.
Thin copper wire: AWG 28 : 2081f/lb (.0005 lb/f); 65.31ohm/1000' (source ARRL Hbk.) about $12 for a 1500 feet spool at Mouser Electronics

Support wire: fishing line type 20+ lb at .0005 lb/f (conservative estimate).

I choosed to evaluate a 200'H x 200'L feet trapezoidal setup, ABCD c.c.w. with base angle at A and D = 60 deg.

XY locations



Sides: AB = CD = 230.94'
Top base: BC = 200'
Bottom base: AD = 430.94' Actual side **line AB** is made of 2 lines of support cable, thus 40 lb strain capacity and unit lenght weight of .001 lb. Reasons for doubling this lenght will be brought later.
Lines BC and **CD** are made of 1 support line and 1 thin copper wire for, again, a unit lenght weight of .001 lb, and strain capacity of 20 lb (only the support wire will bear any constrain, the copper wire only brought along on support wire, for it's antenna function. So only the support wire would be attached to the ground and the balloons.

A cable or wire suspended between to horizontal points, such as points

B and C in the above figure is called a ' Catenary ' in a Mathematical and Mechanical sense. A cable or wire such the ones that links A and B or C and D (sides) is more simply call a ' Flexible Cable Suspended from Two Differents Levels ', at least in the Mechanical engineering texbook I refered. (1952 ed., but the logic still good, even after iron curtain fall ;-))

I will limit here the results resolution of the proposed equations brought, in the case of this problem. I would suggest to those interested by the details of these equations to refer to similar book as mentionned above. They are long, dry and require the use of minimal graphic support, beyond the limit of this media, to have any meaningfull sense. I will just mention they are of the recursive type, that one can resolve with a spreadsheet such as Excel with relative ease.

The resulting figures were found to be:

Top section BC (Catenary):

Actual wire lenght:	202.97'
Sag	15.0'
Tension at B and C	.3508 lb at 16.82 deg with horizontal axe
	.3358 lb along x axis
	.1015 lb along y axis (down)
Side sections AB and CD	
Actual wire lenght:	231.22' *(to compare with 230.94' above, thus very little sag on those)*
Tension at B and C	.7790 lb at 64.47 deg with horizontal axis
	.3358 lb along x axis
	.7030 along y axis (down)
Tension at A and D	.5790 lb at 54.55 deg with horizontal axis
	.3358 lb along x axis
	.4717 along y axis (up)

At B and C vertical down force combine .1015 + .7030 to give .8045 lb which is the vertical upward force (net boyancy) that is required by the balloons, at each site, to maintain the setup aloft at desired location.

The Baloons

As per Archimede's principle, the upward force developped by a balloon is related to the weight of the displaced air by the balloon volume.
Actually the balloons shall be sized to carry this net boyancy (.8045lb) + the weight of the helium (s.g. = .1308 that of air) it carries + the weight of the balloon envelope.
For exp., someone wants, on 500' ASL (at sea level) site put a 200' high (= 700' ASL) balloon able to procure sufficient lift for a .8045 lb payload.

Let suppose that a 40 inches (3.33') diameter of .5 lb envelope weights is available.

Density of air at 500' ASL = .0754 lb/pi3, at 700 ASL = .0750 lb/pi3

Volume = .5236 x d3 = 19.39 pi3

At 700' ASL ; Weight of air = 19.39 x .0750 = 1.454 lb
At 500' ASL ; Weight of helium = 19.39 x .0754 x .1308 = .191 lb

Net boyancy at 700' ASL = 1.454 - .191 - .4 = .863 lb

In this case, 2 of these balloons would be ok for the proposed setup. The top part would be a bit flatter (but, not by much) than calculated

Effects of the wind

In fluid mechanics textbooks, one can find formulas to determine the drag force applied by moving air upon stationary spherical object. (the balloons) :
Fd = Cd p V2 A / 2

Fd : drag force (lb) Cd : drag coefficient (no dimension)
V2 : Velocity of the wind squared (f2/sec2)
A : Center area of the sphere (f2)

Cd = .4 for Re (Reynold number) < 350,000
Cd = .2 for Re > 350,000

Re = D V / u

D diametre of the spere (feet)
V Speed of the wind (feet/sec)
u (nu) Kinematic viscosity of air = ~.00016 (f2/sec)

Below a table that shows the evolution of horizontal pressure (in ponds) induced by various wind velocities (mph) on various balloon diameter

(feet)

Table 1.
Horizontal wind-induced pressures on a spherical balloon: Fd (pounds)

V (mph)	3'	3.33'	4'	5'	6'	Diam
10	.71	.88	.66	1.02	1.5	
15	.83	1.03	1.5	10.7	15.4	
20	1.5	1.83	2.63	4.1	5.9	
30	3.33	4.1	5.9	9.2	13.3	
40	5.9	7.3	10.5	16.5	23.7	

Here, I will limit the analysis to a worst case senario, where I will use only common sense to evaluate resulting strain on support wire AB. Examining the trapezoidal shape of the proposed setup, one can observe that a wind blowing from the left would induce the highest strain on line AB, resulting from the combine effect of the pressure on balloon C 'added' to the one on balloon B through wire BC.
To get these 'exact' values one would have to develop some kind of program that would require to first compute the relative reposi- tionning of each balloon wit varying wind condition and then apply vector computation to get this value. Developping such a program would be time consuming.

Common sense evaluation should suffice to get acceptable ballpark values. One can see that wind blowing from the left will tend 'push' balloon B along a circonference of radius AB, thus going down. At the same time, balloon C would be ' forced ' to travel up along the circonference of radius DC. Those 2 opposite forces would tend to oppose, resulting in reduced displacement of point B as to compare to a single vertical balloon. In any case one can see that balloon B moves at all, the effect will be to reduce the angle at A.

Hence, If one directly adds the horizontal wind induced pressure on B and C balloons (or doubles the value at B) and divide the result by cosine 60 deg, one would get a value of the strain induced in wire AB higher than reality, but still representative for evaluation purpose. I did that to Table 1 above to generate Table 2 below. Let's keep in mind these values are somewhat on the overestimated side.

Table 2.
Estimated side wind-induced strain on wire AB: T (pounds)

V (mph)	3'	3.33'	4'	5'	6'	Diam
10	3.9	4.1	3.1	4.7	6.9	Safe zone
15	3.9	4.8	7.0	10.7	15.4	
20	7.0	8.5	12.2	19.0	27.4	
30	15.4	19.0	27.4	42.7	61.8	Risk zone
40	27.4	33.9	48.8	76.7	110.	

One last table (some useful physical properties of air)

Altitude(ASL) feet	Air pressure lb/f2	Air density lb/f3	Air viscosity f2/sec
0	2116.2	.0765	.000156
1000	2040.9	.0743	.000162
2000	1967.7	.0720	.000164

Well those are my findings. I hope those infos will be of some help for anyone who would like to tackle (or proceed) with the idea of erecting some kind of balloon antenna.

A Skyhook for the '90s

By: Don Daso, K4ZA
Date: May, 1997
Original source: QST May 97, pp. 31-33

Balloons, shaped like small blimps, are in relatively common use for advertising purposes. Such a blimp, of 10- to 12-foot (3.3 to 3.7 m.) length and 3- to 4-foot (0.9 to 1.2 m.) diameter can easily be used to support a vertical antenna for temporary use, such as in Field Days. K4ZA has successfully used them, and in this article, offers the benefit of his experiences for others who might want to follow his example.

A balloon of such size, when filled with helium, will provide a lift of about two pounds, ample for a quarter-wavelength wire on the 160-meter band. The balloon should be tethered in the center of an open space of such size that the perimeter is at least as far in all directions from the tether as the length of the antenna. For safety and good results, one should select an area of relatively flat ground with no buildings, towers, trees, fences, or overhead wires within it.
The balloons are commonly made of polyurethane film of 2- or 3-mil (50- to 75-micron) thickness. The film is relatively strong but must be

protected from puncture. The author suggests laying a 20- by 20-foot (6- by 6-m.) tarpaulin around the tether where the blimp will be inflated, to protect it from twigs or sharp weeds.

The tethering line should be 100-pound test fishing line or 1/8-inch (3-mm.) nylon rope. Stranded wire or aluminum welding wire is light and amply strong for the radiator. A good quality ball-bearing swivel should be used to attach the tether to the balloon's halter, since the wind will cause it to turn, pitch, and roll in all directions. Gloves should always be worn when the blimp is being launched and retrieved. In all cases, one should avoid dangerous situations, such as operating near power lines, in wet or stormy weather, or near airports or areas congested with people. Do not attempt to use heights greater than a simple quarter-wave monopole. Finally, remember that an adequate ground system is at least as important as the antenna; use either four elevated, or up to 60 ground radials for best results.

Two suppliers of these inflatable blimps are:
Toy-Tex Novelty Company, 7315 N. Linder, Skokie, IL 60077, tel (847) 673-6600;
The Blimp Works, 156 Barnes Airship Drive, Statesville, NC 28677, tel (704) 876-6705; web http://www.msmall.com/blimp.

Homebrew Projects, Antenna, RF, Amplifier, Digital Tecniques,
Contest with photos, projects, download...

by **Talino Tribuzio**

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Shortened 160 meter vertical

PI4CC home

BCS
MK II
Additional info from K8GG on the BCS.

Guestbook

Mail PI4CC

Here is a description of shortened 160 meter vertical. The Battle Creek Special is worked out here.

This is a copy from an article in QST December 1974.

Mobile operators and those who reside on property of city-lot size should find the author's treatment of physically shortened 160-meter vertical antennas of considerable interest. Construction details are given here for making his Minooka Special from readily available inexpensive materials. Performance is reported to be excellent with both the mobile and fixed-station versions of his design. Data are given also on basic antenna types in use by some of the leading "top band " DXers.

THE MAIN ACTIVITY at W9UCW for the past 10 years has centered on 160-meter operation. The focus has been on antennas. Much of the study has been directed toward comparing the performance of vertical and horizontal antennas for various objectives. A further study has been made on portable, mobile, and small-lot antennas, including suitable antennas for use at DX locations. Three trips by the author to HK0 and HK1 locations motivated further investigations of portable antennas for use on 160 meters. Improved mobile operation was sought also, seeking solutions to the problerres of efficiency, durability, bandwidth, ease of construction, and cost.

Inspiration for the design offered here resulted in past from the author's survey of antennas in use by 160 meter operators, worldwide. An extensive report was compited. It filled a loose-

leaf notebook! A boil-down of the most useful information resulted in an 11-page report which was made available to numerous amateurs. Subsequently, the writer was encouraged to submit some of the data for publication. Approximately 60 operators were polled, The following information from that inqury should be of interest to those who are curious about "preferred"antennas for topband use. Question if you could put up any antenna for 160, what would it be? Result: 60% favored verticals, 30% said horizontals, and 10%. indicated mixedfeelings. (The term "vertical includes various configurations 1/4, 1/2, and 5/8 wavelength, vertical arrays, and inverted Ls.) Questions 2: "Comparing antennas that an average ham could build, do you prefer verticals or horizontals for 160? Result: 70% said vertical, 17% favored horizontal, 5% inverted L, 2% horizontal and vertical 2% inverted V and 3% had no opinion. Reason; given for the responses were, "Because of signal comparisons and past experience. A backyard-compatible vertical is more effective than a back-yard-compatible horizontal." Question 3: "Do you operate mobile on 160? If so, what is the antenna used?" Result: 3 used base-loaded verticals 8 used center-loaded ones and one employed a Heliwhip. Some of the operators used capacitance hats. Other questions in the surrey dealt with receiving antennas, types of soil and terrain, besides requesting complete details of the present transmitting antenna.

The Minooka Special

The following is a description of an effective vertical antenna for 160 meters, designed with these objectives in mind.

1) Highly effective for 160-meter DX and local work. 2) Easy to build and adjust. 3) Very economical. 4) Fits neatly into back yard. 5) Reasonable bandwidth. 6) Good for portable and I)Xpedition work. 7) Can be scaled down for mobile operation.

The resultant antenna (Fig. 1) is top loaded inductively and can be built by anyone from readily available material. Only a dip meter and SWR indicator are needed for tune-up Man versions were built and tested, ranging from 7- foot mobile types to 60-foot backyard or DXpedition models. They have been used with good results from 20 foot- wide backyards in cluttered Chicago,

to vast beaches on Carribean islands, and in South America. The fixed-tuned bandwidths vary from 10 kHz for mobile versions, to 50 kHz for the larger fixed-station models (SWR 2:1 or better).

Fig. 1 Electrical details of the Minooka special.
Table 2 gives specific information concerning dimetions X, Y and Z L2 may vary in size from 1 to 20 turns, and L3 will contain between 5 and 10 turns, typically. L2 and L3 are made from No.18 wire spaced 1/8 inch between turns. The coil diameter is 1-1/2 inches. Refer to text for tuning instructions.

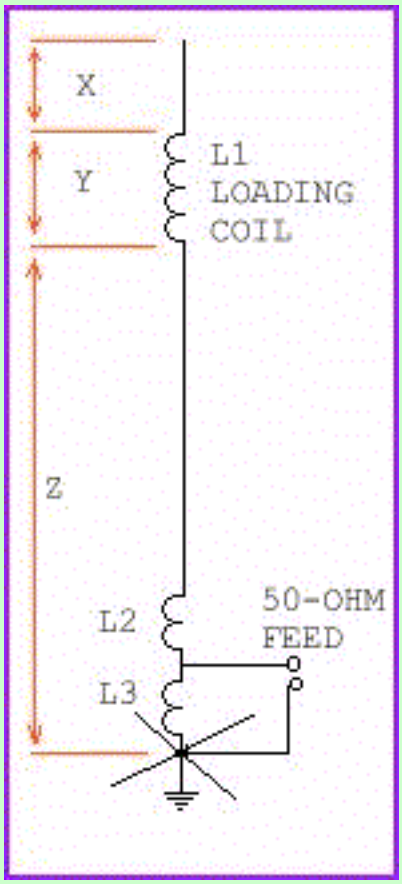


Table 2

X	5 ft	8 ft	4 ft	4 ft	19 ft	3 ft 3 in
Y	2 ft	15 in	3 ft 6 in	4 ft	11 ft	3 ft
Z	*	*	*	*	*	*
Wire size	No 20	No 19	No 18	No 16	No 19	No 22

* mean as long as possible

Construction

The physical layout of the antenna is centered around the use of 3/4-inch-diameter rigid PVC water pipe. This sturdy tubing has an outside diameter of 1.038 inches. The inner diameter provides an "interference fit" for 1/2-inch EMT conduit (thinwall). A 3/8-24 nut is driven into a 5-inch length of conduit, then aligned and braized into position. Next, the conduit is taper-ground and polished on the opposite end, then driven into the PVC tubing (see table of dimensions).

In the construction of mobile antennas another piece of conduit is driven into the bottom of the PVC coil form and a 3/8 - 24 bolt is brazed into the bottom end The bolt will mate with standard mobile antenna mounts. For larger versions of the antenna the PVC material can be mounted in TV wasting, or whatever. A standard 8-foot stainlesssteel whip between 3 and 8 feet in length can be screwed into the top section. For home-installed versions of the antenna a three- or four-foot piece of thin-wall tubing can be used in place of the whip. This will save on the cost of materials, and will eliminate the need to have brazing done. The coil wire should be soldered to the conduit to assure a good electrical connection.

Adjustment

Pick a set of dimensions from the table which suits your application, but add a few inches more of coil turns (all turns close wound) than are recommended This will allow leeway for pruning the system to resonance.

To simplify adjustment it is suggested that the system be assembled first with only the coil and top section (no base section). Place the antenna where it is in the clear (on the car or fixed-station site), and where it can be tuned against the proposed ground system - car body or ground radials.

A three-turn link should be connected temporarily between the lower end of the coil and the ground system. This will permit rough tuning of the system to resonance by inserting a dipper

coil into the link and adjusting the coil turns on the antenna until a dip is noted in the desired part of the band. Upon completion of the pruning the constructor can, if he wishes, cover the coil with weatherproof tape or shrink tubing. The antenna should be tuned for roughly 2000 kHz if the entire band is to be used. Adjust the resonance for 1850 kHz if only the low segment will be utilized.

Erect the antenna with all of its parts - coil, top and bottom sections - and insert inductors L2 and L3 of Fig. 3 as shown. With L3 temporarily out of the system, adjust L2 for the lowest value of SWR obtainable at the desired operating frequency. Then, place L3 back in the circuit and adjust it for an SWR reading of 1. Addition of the base section later on will not affect the resonant frequency of the overall system materially, provided the base section does not exceed, say, 60 feet.

The foregoing steps are used also in adjusting the mobile version of the antenna. However, because of the small size it is possible to adjust the antenna in completely assembled form. Only L3 of Fig. 3 is needed for mobile antennas. The main coil, L1, is adjusted for resonance, then L3 is set for lowest SWR. L3 can be mounted inside the trunk or under the bumper in a weatherproof enclosure.

For fixed-station operation it is recommended that a good ground system be employed. One should use at least 10 radials oft say, No. 18 or larger wire, 10 to 50 feet in length. If you can manage 40 radials, 60 feet in length, all the better.

Concluding Comments

Three fixed-station versions of the Minooka Special have been tested and used at W9UCW. Each was compared against the regular antenna, which is a one-quarter wavelength vertical (130 feet high), and operates against a radial system that contains 12,000 feet of wire. The short verticals were always inferior to the big antenna by approximately 5 dB.

Using version No. 4 from the table (4-foot base section), the author worked Wls through W6s from the mobile setup. The rig

was an HW-18 transceiver. Contacts were also made with VP9 and KV4 stations from the car.

While DXing from San Andres (HK0) reports were received of signal strengths exceeding S9 plus 20 dB from Maine to Washington. Good reports were received from Europe also.

Because the loading coil acts as an rf choke at 3.5 MHz and above, several versions of the antenna have been used successfully from 160 through 20 meters with an appropriate L-network matching section installed at the base of the system.

Acknowledgements

The author wishes to express his gratitude to all who helped make this article possible particularly K9SKX, WA9EYY, and W9YYS, who assisted with the initial experiments on the Minooka Special. The participants in the antenna survey (a list too large to include here) deserve considerable thanks for their help as well.

The Battle Creek Special is worked out here.

The OK1RR DX & Contesting Page is moved!

The new URL is:

http://www.ok1rr.com

You will be redirected in 10 seconds.
If not click on the link above.

Many thanks to Al Waller, K3TKJ for his long and outstanding service.

OK1RR DX & Contesting Page

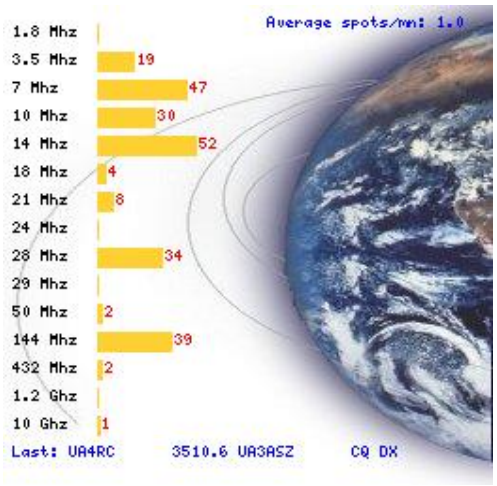
Search this site:

Enter phrase:

Statistics

You are visitor # since May 18, 2004 (in this domain)

The bands just now



Welcome

This page is devoted exclusively to Amateur radio. You may find here some stuff concerning DX, CW and contesting.

If you have any great suggestions, feel free to **email me,** and I'll probably feel free to ignore you.

How to use the 'The bands just now' chart?

This bar chart is based on the 250 lasts spots from the OH2AQ web cluster (fantastic tool!).

For each band a bar chart is drawn proportionally to the number of spotted DX informations. The absolute number of spots is noted in blue at the end of each bar. To see more spots from a particular band, click on the band indicator.

The last spot posted on the cluster is noted in blue at the bottom of the graph.

The Average spot/minute indicator is a calculated stamp, based on the 25 lasts spots.

by F5LEN

Data Sheets of 3225+ components

Download in PDF format, many hard to find ones there! The download is limited to 10 items per day. Don't exceed this number, please, unless your IP would fail into banned list!

*courtesy of **AMA-CB SERVICE** OK1DNH*

PLC (Power Line Carrier) or PLT (Power Line Technology) - the end of Short Wave Bands?

They will use HF and VHF broadband systems. The outdoor power wires will serve as antennas, typically producing directivity and hugely strong increases in background noise for ham reception. ARRL did a study and included it in Comments in the FCC docket. Also, consumers will have equipment potentially susceptible to the ham signals. FCC rejected access to the 136 kHz band because of the risk of interference at 3mV/m, whereas the ham signals in HF/VHF will be at 30mV/m. ... **Download and read the set of basic documents.**

Notes on contesting software

Tired of non standard hardware devices or software solution? Writing a contesting program? **Read my notes** on contesting software and avoid common mistakes which I collected within past 12 years.

Super Check Partial Databases Download

- **All callsigns**
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- **W/VE**
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Master Database with Club Memberships

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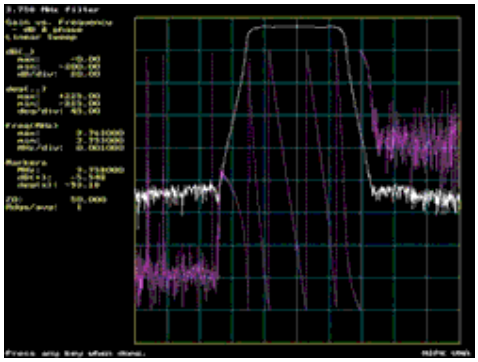
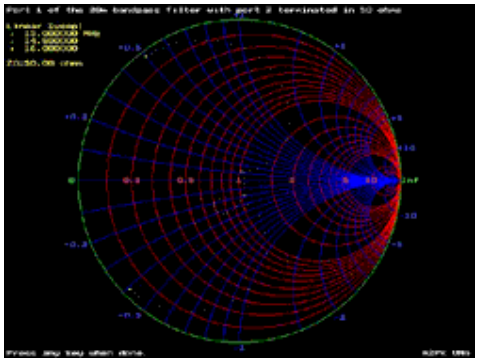
SPAR Society For The Preservation Of Amateur Radio

VNA - The Vector Network Analyzer by N2PK

Not satisfied with your MFJ-259B, Autek RF-1 or AEA CIA HF? Impossible to spend some 2 kilobucks for an advanced vector impedance analyzer? Do you want a highly interesting home brew project? Then visit **Greg Ordy's (W8WWV)** and **Paul Kiciak's (N2PK)** web sites and learn more about the VNA - Vector Network Analyzer built and developed by N2PK. This is a homebrew VNA capable of both transmission and reflection measurements from 0.05 to 60 MHz, with about 0.035 Hz frequency resolution and over **110 dB of dynamic range**. Its transmission measurement capabilities include gain/loss magnitude, phase, and group delay.





(click on images to enlarge)

Its reflection measurement capabilities include complex impedance & admittance, complex reflection coefficient, VSWR, and return loss. Unlike other impedance measuring instruments that infer the sign of the reactance (sometimes incorrectly) from impedance trends with frequency, a VNA is able to make this determination from data at a single frequency. This is a direct result of measuring the phase as well as the magnitude of an RF signal at each test frequency.

With optional narrowband extensions, this VNA can function through 500 MHz with some degradation in performance. Its custom software operates on DOS and Windows based IBM compatible PCs and communicates with the VNA through a parallel port. It is aimed at the serious experimenter with, at least, a basic understanding of transmission lines.

A sad story of the stolen name of the HSC

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1437

A1 - OPERATOR CLUB

731

61

130

1731

151

1994

MARCONISTA

223

OKDXC
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24

OK DX Foundation

QSL Manager Database available for download again!

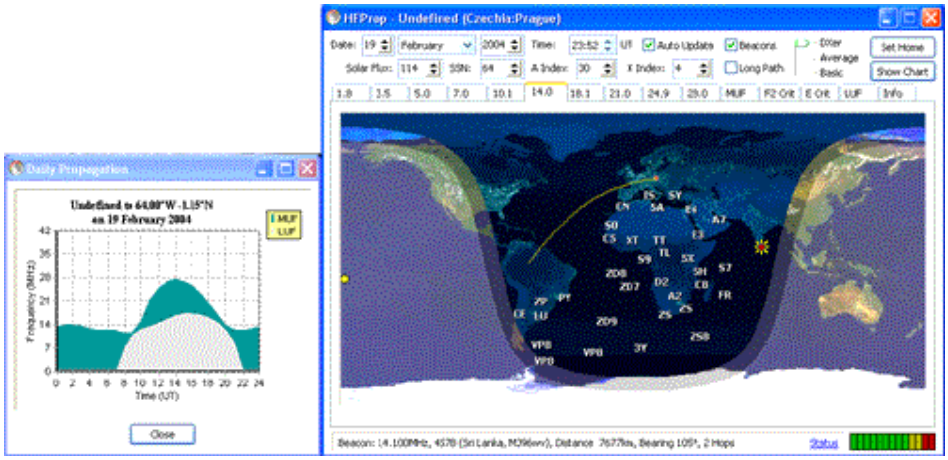
The new version is still in process which take some time. My apologies for this inconvenience. If you want to read some news, **click here**.

Software news

Software is like sex: it's better when it's free.

Linus Torvalds

- **HFProp** by Julian Moss, G4ILO is very nice hard-to-beat software. It employs Solar Flux or SSN value and both A and K displaying a particular propagation prediction chart to any point on any band at a glance. Many other programs are hard to use because it cost a significant effort to prepare a prediction to single point. HFProp combines a powerful beacon tracker with very good propagation calculator dusplying the results on a Mercator world map. Julian says that it employs "a quite simplistic algorithm" but I can witness that it works at least in the same way as many of the highly priced programs. You can also directly compare the results with beacon availability, therefore confirm the prediction accuracy.



- **CT by K1EA** (DOS version) is **now free!** There are many new features including paddle support. Look for regular updates at the CT home site.

- **Swisslog** by Walter Baur, HB9BJS for DOS is another addition to high value **free** programs. The latest DOS version 3D8a is probably the best logger for DOS which beats many known and expensive Windows logging programs.

- **4nec2** by Arie - a top grade **FREE** antenna analysis software based on true NEC2. Generates very nice color 3D patterm models, Smith chart, can optimize your antenna design and works up to 11000 segments, based on the RAM capability of your machine. The only disadvantage is that it uses NEC cards, you need a bit of experience with this ancient method. I believe this is the BEST antenna software!

- **NEC2 for MMANA** by Dimitry Fedorov, UA3AVR- a supplement to the famous MININEC implementation allowing NEC2 antenna analysis up to 3000 segments. Calculates also structures made of dielectric coated wires. Works with .maa files which can be prepared in the MMANA. There is a way to convert a MMANA file to the NEC card file (see above), simply rename the **Indata.txo** file generated by this program to a file with .nec suffix. An ideal companion to MMANA and 4nec2!

- **VUSC** (**V**HF - **U**HF - **S**HF **C**ontest) by OK1DIX/N1GA is a TR-like contesting freeware with many advanced features. Visit the **author's web site** to obtain the most fresh version. The new **version** features:

  - Both modes on-line (during the contest) and off-line (evaluation after the contest) possible
  - Free log data editor
  - Multiband support (up to 8 VHF/UHF bands simultaneously on one computer)
  - Very high memory capacity (several thousands QSOs, all databases in RAM)
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 # 47

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Drag the text with mouse or wait until text scrolling.

Upcoming operations

```
01/01-31/12      IOTA 2004
till  ??         9A/Z32FD: Pag Island (EU-170)
till  ??         RK3DZJ/1 and R3ARS/1: EU-147 and EU-066
till  21/08      5X2A: Uganda *  by K4ZLE
till  21/08      TM3ID and TM6ANV: special stations (France)
till  22/08      7Q7HB: Malawi * by G0JMU
till  22/08      SM0DTK/1: Gotland Island (EU-020)
till  23/08      EI/KC8FS/p: Ireland
till  23/08      MM0KAL, MM0XAU, MM5PSL: Sumburgh Head Lighthouse
till  24/08      F/M0ARK/P: Re Island (EU-032)
till  24/08      PJ: Bonaire (SA-006) * by I4ALU
till  25/08      VY0CQ * by VE3ZCO
till  27/08      5U7DX: Niger * by PA5M
till  27/08      EO60FO: special call (Ukraine)
till  27/08      HH4/K4QD: Haiti (NA-096)
till  27/08      OZ/DL2VFR: Bornholm Island (EU-030)
till  29/08      F4AJQ/p: Noirmoutier Island (EU-064)
till  29/08      SY2004PHG: Greece * by SV1AGU
till  31/08      4O6100BB: special station (Montenegro)
till  31/08      ES1924: special event callsigns
till  31/08      S79OA: Mahe Island (AF-024) + Alphonse Island (AF-033)
till  August     HE5IBC: special call (Switzerland)
till  August     IR8M: Calabria region islands * by MDXC
till  August     VY2MM: Prince Edward Island (NA-029) * by W3KHZ
till  01/09      9A2CY/P: Cres Island (EU-136)
till  01/09      J42004A: Skopelos (EU-072) & Athens * by M0BBB
till  04/09      SV5/SM8C: Kalymnos (EU-001) * by SM0CMH
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 # 355

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 # 186

RCC

and others

till	07/09	VO2/K2FRD: Labrador (Zone 2) * by K2FRD
till	15/09	J42004: special prefix
till	15/09	RA3XR/0 and UA3YH/0: Dikson Island (AS-005)
till	September	3D2EA: Viti Levu (OC-016) * by EC3ADC
till	September	ES7NY/2: Naissaar Island (EU-149)
till	September	FO5RN/p: Tahiti (OC-046), French Polynesia * by F5MJV
till	September	II0P: Sardinia (EU-024) * by IZ1EPM
till	September	YA0Y: Afghanistan * by DL5SE
till	September	YJ0XX: Vanuatu * by N5XX
till	31/10	SV0XAN/5: Dodecanese (EU-001) * by IK2WZD
till	04/11	II3T: Special event call * by ARI Trieste
till	15/11	SX2004 and SY2004: special prefixes
till	November	EL/EI5IF: Liberia
till	31/12	9A80ADE, 9A80Z, 9A80ABD: special event stations
till	31/12	HA2004EU: Special event call
till	31/12	HB75A: Switzerland (USKA 75th Anniversary)
till	31/12	HS72B: special call and licence (Thailand)
till	31/12	NL7AU: Upper Matecombe Key (NA-062)
till	31/12	OE80XRW: special event station
till	31/12	SG1RK: special event call (EU-020)
till	31/12	W1AW/90: ARRL's 90th anniversary
till	December	HF0QF and HF0POL: "Arctowski" Base (So. Shetlands)
till	December	OX2KAN: Special event station
till	December	VK0DX: Davis Station (Antarctica) * by VK4LL
till	December	VQ9LA: Diego Garcia (AF-006)
till	December	WL7CPA: Unalaska Island (NA-059)
till	December	YI9KT: Iraq * by SP8HKT
till	December	YI9MC: Iraq * by KC4MC
till	January	HA200CVM: special station (Hungary)
till	March 2006	5H3HK: Tanzania * by JE3MAS
till	March 2006	ZD8I: Ascension Island * by G8WVW
16/08-30/08		IG9/IK2MLR: Linosa (AF-019)
18/08-16/09		ZX35O: special call * by PT2OP
19/11-21/11		SEANET Convention
20/08-22/08		4Z4DX: Jaffa Lighthouse
20/08-22/08		CS0RCL/LH: Berlenga Island (EU-145) & LH
20/08-22/08		DD6VSF: Ruegen Island (EU-057)
20/08-29/08		ER60EM: special station
20/08-22/08		F6HKS/p: Leucate LH
20/08-25/08		LA5UKA/p and LA9VDA/p: Gossen Island (EU-056)
20/08-22/08		OI3W: Ronnskar Island (EU-097) & LH
20/08-28/08		V47UY: Nevis Island (NA-104) * by KJ4UY
21/08-22/08		CU6/CU3AA and CU6/CU3EJ: Ponta da Ilha LH (EU-175)
21/08-22/08		CU6/CU3AA and CU6/CU3EJ: Sao Mateus LH (EU-175)
21/08-22/08		CX1TA: Cabo Santa Maria LH
21/08-22/08		ED8LGP: Barlovento LH (AF-004)
21/08-22/08		F5IRC: Antifer LH
21/08-22/08		F6KUM: Cap de l'Ailly LH
21/08-22/08		GB2DL: Dunollie LH
21/08-22/08		GB2ELH: Eshaness Lighthouse
21/08-22/08		GB2GNL: Girdle Ness LH
21/08-22/08		GB2LBN: Barns Ness LH
21/08-22/08		GB2LT: Turnberry LH
21/08-22/08		GB2MSL: Kinnaird Head Old LH
21/08-22/08		GB2NCL: North Carr Lightvessel
21/08-22/08		GB2RRL: Rubha Reigh Rua Reidh LH
21/08-22/08		GB2SHL: Stoer Head LH
21/08-22/08		GM3TKV/p: Kingston Beacon
21/08-22/08		HP2L: Gamboa LH
21/08-29/08		IA5/IK5WOB and IA5/IK5FTL: Isola d'Elba (EU-028)
21/08-10/09		IF9/I5RDF: Marettimo Island (EU-054)
21/08-22/08		II3T: Vittoria LH

21/08-22/08	IU1L: La Lanterna Lighthouse
21/08-22/08	K8E: Eagle Harbor LH
21/08-22/08	LT7W: Golfo Nuevo LH
21/08-22/08	LU2DT/LH: Punta Mogotes LH
21/08-22/08	LU: Atalaya LH * by Grupo Oeste
21/08-22/08	Homecall/D: Recalada LH * by LUs
21/08-22/08	Homecall/D: El Rincon LH * by LUs
21/08-22/08	Homecall/D: Segunda Barranca LH * by LUs
21/08-22/08	homecall/V: San Matia LH * by LUs
21/08-22/08	homecall/V: Villarino LH * by LUs
21/08-22/08	MM0MWW: Hoxa Head LH
21/08-22/08	MM1HMV: Toward Point LH
21/08-22/08	MM3STM: Ardrossan Pier Head
21/08-22/08	OH/YL2FB: Soderskar LH
21/08-22/08	PI4LDN: Noordwijk LH
21/08	PI4VPO/LT: Stenen Baken LH
21/08	PY7XC/p: Itamaraca Is. (SA-026)
21/08-22/08	TF1IRA: Knarraros LH
21/08-22/08	TO0FAR: Sainte Suzane LH (Reunion Island, AF-016)
21/08-22/08	UU9JWM/P: Khersonesskiy LH
21/08-22/08	VI2MI: Montague Island (OC-223)
21/08-22/08	W2T: Tucker's Island Lighthouse
21/08-22/08	ZV7AA: Natal LH
21/08-22/08	International Lighthouse/Lightship Weekend
21/08	Japan DX Meeting 2004
21/08-22/08	SARTG WW RTTY Contest
21/08-22/08	Seanet Contest
22/08-25/08	2M0NJW and GM4RQI: St. Kilda (EU-059)
22/08-06/09	9A: EU-136 * by IN3YGW and IN3DEI
22/08	9H4GRS/P: Jordan's Lighthouse
22/08	CQ0RLH & CT1GPX/LH: Cabo Raso Lighthouse
23/08-31/08	AT0RI: Pamban Island (AS-???)
24/08-31/08	XX9: Coloane Island (AS-075) * by JA0SC
25/08-22/09	GB6LOP: special station (Liberation of Paris)
26/08-07/09	8Q7JF & 8Q7GA: Maldives (AS-013) * by DL7JAN & DL3GA
27/08-29/08	BV9A: P'enghu Island (AS-103)
27/08-01/09	GB5FI: Flatholm Island (EU-124)
27/08-29/08	R1MVI: Malyj Vysotsky (EU-117)
28/08-04/09	HB0/HA4XG: Liechtenstein
28/08-29/08	SCC RTTY Championship
28/08-29/08	YO DX HF Contest
29/08-12/09	K3GV/VY2: Prince Edward Island (NA-029)
August	UE9OWM/9: Altay-Himalaya Transasian expedition
August	ZK3: Tokelau Islands * by EC3ADC
01/09-08/09	GB4IOM and GB4SPT: Isle of Man (EU-116)
02/09-12/09	T98LBC: Bosnia & Hercegovina * by DL7AFS, DJ7ZG, Z31GX
04/09-05/09	IQ7AF: Lecce - Salento DX Team Meeting
04/09-05/09	TM5BDM: special station (France)
04/09-18/09	TM8CDX: special station (France)
04/09-05/09	All Asian DX Contest Phone
08/09-18/09	SV8/ON4BB, SV8/ON5JE, SV8/ON5KH: Naxos (EU-067)
09/09-23/09	7Q7CE: Malawi * by IN3VZE
09/10-23/10	VK9L: Lord Howe (OC-004) * by Oceania DX Group
10/09-18/09	FP/NN9K: Miquelon Island (NA-032)
11/09-12/09	CIS DX Contest RTTY
11/09-12/09	WAEDC - Worked All Europe DX Contest SSB
18/09-20/09	FO/IT9YRE, FO/I1SNW, FO/IT9EJW: Hereheretue (OC-052)
18/09-21/09	LX/ON6QX, LX/ON4LO, LX/ON6UM: Luxemburg
18/09-19/09	MIA Contest (www.mdxc.org)
18/09-19/09	Scandinavian Activity Contest CW
19/09-21/09	SV8/ON4BB, SV8/ON5JE, SV8/ON5KH: Mykonos (EU-067)
19/09	PSK31 Contest

20/09-30/09	CT9R: Madeira (AF-014) * by EAs
21/09-28/09	GB2LI: Lundy Island (EU-120)
22/09-29/09	PJ4/K9MDO, PJ4/N2WB, PJ4/W9ILY: Bonaire (SA-006)
23/09-26/09	OJ0YC: Market Reef (EU-053)
23/09-02/10	VP5/AH6HY: Turks Islands (NA-003)
24/09-27/09	W4D: Dauphin Island (NA-213)
25/09-26/09	Scandinavian Activity Contest SSB
25/09-26/09	XX Italian HF-DX Convention (Bologna)
18/10-23/10	V6O: Pulap Island (OC-155) * by W5BOS
22/10-24/10	RSGB International HF & IOTA Convention
23/10-02/11	FP/VE7SV: St. Pierre et Miquelon (NA-032)
25/10-26/10	CQ World Wide DX Contest RTTY
26/10-02/11	J75WX,J79AA,J79LR,J79CM,J79VL,J75J: Dominica (NA-101)
30/10-31/10	VK4WWI/8: Elcho Island (OC-185) * by PA3EXX
October-November	3B8MM: Mauritius (AF-049) * by DL6UAA
October	KH7K: Kure Atoll (OC-020)
October	TX0: Chesterfield Is. (OC-176) * by DL5NAM and others
01/11-02/11	VK4WWI/8: North Island (OC-198) * by PA3EXX
03/11-04/11	VK4WWI: Sweers Island (OC-227) * by PA3EXX
08/11-12/11	VK4WWI/p: Marion Reef (OC-???) * by PA3EXX
22/11-09/12	VK9XG: Christmas Isl. (OC-002) * by W0YG
28/11	3D2FI: Viti Levu (OC-016), Fiji * by G0UIH
29/11-05/12	3D2FI: Nacula Island (OC-156), Fiji * by G0UIH
November	PT2GTI: "Comandante Ferraz" Station, South Shetlands
06/12	3D2FI: Viti Levu (OC-016), Fiji * by G0UIH
07/12-10/12	3D2FI: Beachcomber Island (OC-121) , Fiji * by G0UIH
11/12	3D2FI: Viti Levu (OC-016), Fiji * by G0UIH
21/01-04/02	3Y0X: Peter I Island (AN-004)

source: 425 DX News

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Best viewed with any Java capable browser and 1024x768 resolution.

The WE6W Hi-Voltage 160 Meter QRP Loop Ant.
Homebrewed Link-coupled experimental tuner.

Standard Double Bazooka design!
Std. Bazooka I built.
Std. Bazooka compared to my dipole
Folded Bazooka! 300 Ohm feed!

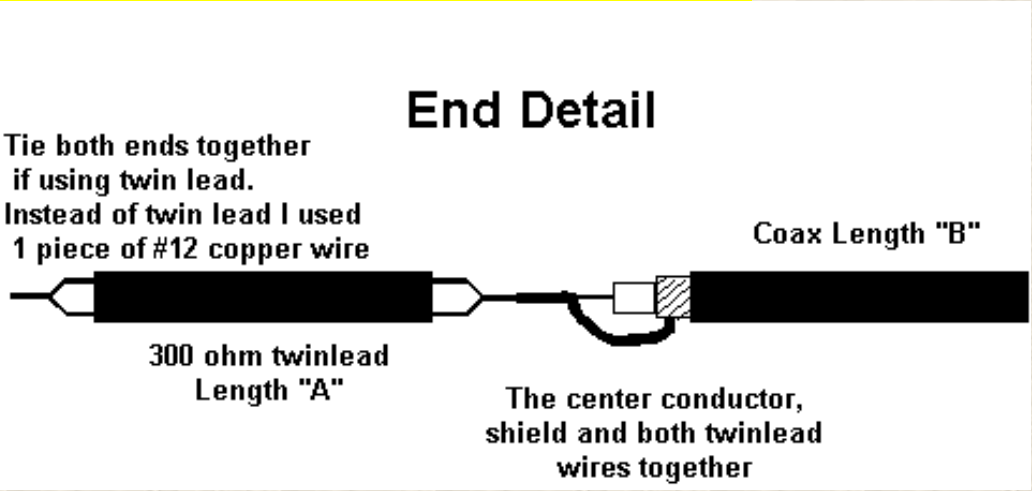
WE6W Field Day Antena design!

Build a Sterba Curtain Antenna!
Notes:Sterba Beamwidth/Gain

St. Louis Vertical
Broadband Antennas
The A613 broadband bow-tie
W4RNL L. B. Cebik:Antenna articles/Software!

Back to we6w's Home Page

THE DOUBLE BAZOOKA



HAM UNIVERSE

SET YOUR SIGHTS ON BETTER QSO'S

Notes on Assembly:

On the cable ends you do not need to use twin lead. You can make these antennas using a single piece of 12 gauge copper wire for each end,, you can also use ladder line etc. I would advise using heavy end wires for strength purposes. To prevent the joints from breaking especially for long lengths, I picked up some 1/2 inch PVC pipe and end caps. I cut the PVC so it would overlap a few inches on the joints. I then glued the end caps on. Place the PVC between the joints and secure with regular screw hose clamps. Put the clamps close to the end caps and it will give a strong joint. Do this for the center also.

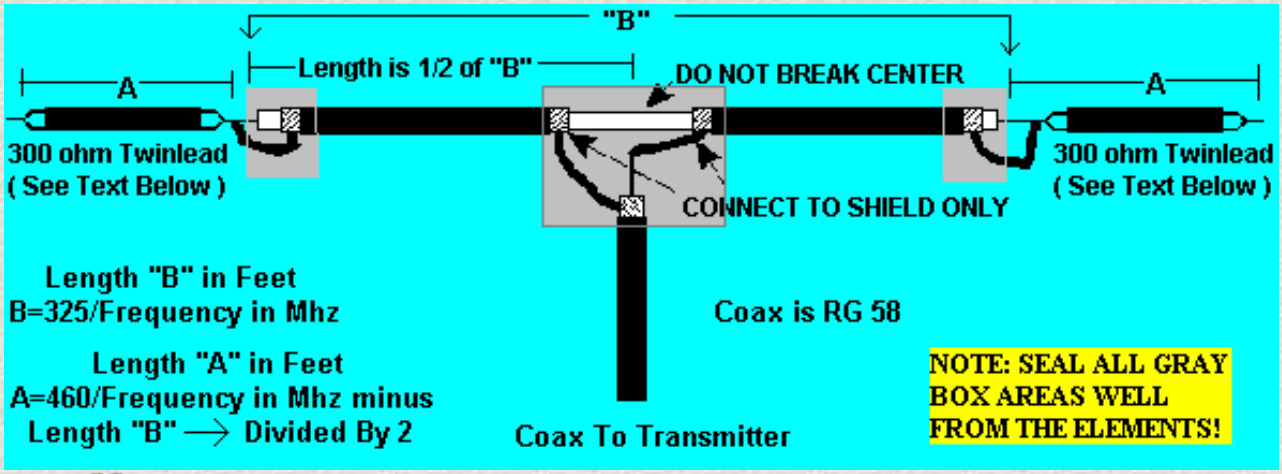
About The Antenna

The Double Bazooka Dipole is a very efficient **single band antenna** which is very quite,and does not require the use of a balun. This antenna consists of coax (RG58) with the shield split at the center and the feedline attached to the open ends. **Do not break the center conductor.**
With the feedline attached directly to the two open ends this acts as a half wave dipole along with the open wire end sections. This antenna can be cut for any operating frequency and is broad banded. It can be mounted as a flat top or an inverted vee and will handle the legal limit.
As with all antenna projects, you may have to tune length for best swr and select materials that are easily obtainable.
EXPERIMENT! EXPERIMENT! EXPERIMENT!.

AN EXAMPLE ANTENNA

75 METER GENERAL BAND CENTER (3.925 mhz)
LENGTH "B" = 325/freqmhz = 82.8 FEET (aprox 82 feet 9 1/2 inches)
LENGTH "A" = 460/freqmhz = 117.20 feet 2 1/2 inches minus length "b" (82.8feet) = 34.4 feet divided by 2 = 17.2 feet
TOTAL ANTENNA LENGTH = 117.2 feet 2 1/2 inches (aprox)

BACK TO ANTENNA LAB



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KB1GW's collection of Beverage Antenna Information

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Sources of Gear

Airline Web pages

TS-850 Mods

Introduction

The Beverage (or "wave") antenna was invented in the early 1920s by Dr. Harold H. Beverage. It was first discussed in a paper titled "The Wave Antenna - A New Type of Highly Directive Antenna" written by Beverage, Chester W. Rice and and Edward W. Kellogg for the journal of the American Institute of Electrical Engineers (Volume 42, 1923). The paper discusses testing longwave antennas (7,000 to 25,000 meters; 12-43 kHz) that were 7 miles (11 km) long. This work was done at Riverhead, Long Island, NY, and mentions "shortwave" tests around 450 meters (665 kHz) as a practical upper limit in subsequent experiments. While others have since written about the antenna, if you can find a reprint of this original work in a research library, you'll find the paper is fascinating reading.

In 1938, the Radio Institute of America presented Dr. Beverage with its Armstrong Medal for his work in the development of antenna systems. The Beverage antenna, the citation said, was "the precursor of wave antennas of all types." Dr. Harold Henry Beverage, Stony Brook, NY, USA, passed away on January 27, 1993 (at age 99).

A classic Beverage receiving antenna requires a lot of space. It is a long wire, one or more wavelengths long, mounted near to the ground and oriented in the direction of the desired reception. A nominal 9:1 balun is required at the juncture of the wire and 50- or 75-Ohm coaxial feedline.The far end is terminated with a nominal 600-ohm resistance. (When available land will not permit the installation of a "full length" Beverage, some people install "short" Beverages, ranging in length from about 300 feet up to 600 feet or so.)

The Beverage antenna is highly directional, responsive to low-angle signals, has little noise pick-up, and produces excellent signal to noise ratios. Some say the frequency range suitable for Beverage antennas ranges (from an Amateur Radio viewpoint) from 1.8 MHz on up to about 7 MHz or so...
However, consider the following from Frank, W3LPL:

Subject: Beverage antennas effective on entire HF range
From: Frank Donovan (donovanf@jekyll.sgate.com)
Date: October 22, 1995
Organization: (Usenet's) rec.radio.shortwave

Properly designed Beverage receiving antennas are very effective across the entire HF frequency range. At the W3LPL DX contest station we use Beverages from 1.8 to 14 MHz, and during the sunspot maximum we used them up to 28 MHz!
Beverage arrays (multiple Beverages designed to operate as a phased array) are even more effective on HF. I've seen Beverage arrays with as many as 128 individual Beverage elements, each 220 feet long and 4 feet high.

"How To" articles:

An original article by KB1GW: "Beverage Notes"
(Note: This article originally appeared in the YCCC newsletter: "Scuttlebutt;" April 1997.)

And, another Beverage article by KB1GW (with guest, K4VX):
"More Beverage Notes" -- Copyright: August, 1997.
(Note: This article originally appeared in the YCCC newsletter: "Scuttlebutt;" October 1997.)

An article by K5ZD: "A $50 Beverage"
(Note: This article originally appeared in the YCCC newsletter: "Scuttlebutt;" October 1995 .)

A link: "W3LPL on Beverages"
"Frank Donavan, W3LPL, discusses beverages on the CQ-Contest reflector.
This is a reprint of Frank's message," along with a nice picture of W3LPL's QSL card!

Contest-reflector postings from W3LPL on Beverages:
Close-spaced Beverages

Phased Pair of Beverages

Terminating / Sloping Ends of Beverages

Minimum and Optimum Lengths for Beverages

New! W3LPL Receiving Filters...
"These filters have several applications:

Preventing interference -- or receiver damage -- from strong nearby signals in a multi-transmitter contest station (or field day, etc.), or to reduce interference in the 160M band from nearby AM broadcast stations operating below 1500 kHz."

"Sharing an antenna among up to four receivers on separate ham bands; specifically a diplexer (constructed from two 'LPL filters) is used when you want two receivers -- on separate bands -- to share a common antenna, any multi-radio contest station can benefit from sharing Beverages among radios with less loss than the typical 3 dB power splitter."

A link to a 1968 interview with Dr. Harold H. Beverage
"Dr. Beverage was formerly director of radio research for RCA Laboratories and also was vice-president in charge of research and development for RCA Communications. During the interview, Dr. H.O. Peterson came in and joined the interview. Dr. Peterson was formerly in charge of the Reception Laboratory at Riverhead. The first RCA Laboratory was established by Beverage and Carter in a tent at Riverhead [NY] in 1919. The interview describes the events leading up to the formation of RCA in the autumn of 1919, the invention of the wave antenna, and first diversity reception. It also describes Dr. Hansel's development of the first crystal controlled transmitter and the first 15 meter transmitter. Some of the personal characteristics of Marconi, Dr. Alexanderson, Roy Wegan, Dr. Hansel, Major Armstrong, President Wilson, and others are discussed."

And, a link to a 1992 interview with Dr. Beverage.

Receive antenna's--related resources:

Hardware:

A link to: Oak Ridge Radio, in Littleton, MA
"Manufacturer of quality radio receiving equipment for DXing the shortwave and mediumwave bands." Including an analog opto-isolator to make remote-controlled variable Beverage-terminating resistors. And "RCT Beverage antennas:" Complete remote-controlled-termination Beverage antenna kits.

Industrial Communication Engineers, LTD.
PO Box 18495
Indianapois, IN 46218-0495
Tel. 800-ICE-COMM (800-423-2666)
Main Office 317-545-5412

Cust Serv (parts) 317-547-1398
Fax 317-545-9645
Telex I.C.E. 27-440

ICE offers a Model 180A matching box for $39 (plus $4.50 for shipping). The 180A has taps to select 50 or 75-Ohm coax feedlines; and taps to match 300/450/600 or 800-Ohm Beverage antenna loads. The 180A has dc blocking capacitors and a gas-discharge lightning protection system. ICE also sells a Model 181A ($39), which allows you to apply a dc voltage into your Beverage for remote switching. Like the 180A, the '181A has a gas-discharge protection system. Finally, they offer a Model 185A "resistive load" to terminate your Beverage with. It has same high-impedance taps as the Model 180A and it costs $34. These units are rated for 10 W of continuous RF and 100 W on peaks. (I was told that these ratings are not specified for transmitting into the boxes. Rather, they are what the boxes can withstand when your Beverage picks up energy from nearby transmitting antennas.) All of these boxes are made of 1/8-inch extruded aluminum (milled and tapped). And, if you're looking to buy American, they're all made in the USA.

A link to: ByteMark Corporation in Orlando, FL; tel: (800) 679-3184 (They sell Amidon stuff.)

Among other things, they offer high power ununs designed by Dr. Jerry Sevick, W2FMI...

Features:

- Broad Bandwidth (1 MHz up to 50 MHz)
- Low energy loss (<0.2 dB)
- Aluminum casing with SO-239 connectors on inputs and outputs
- Dimensions: 4 inches x 2 inches x 2.8 inches (same for all models)

Typical Applications: Matching unbalanced 50 ohm coaxial cable to:

- Ground fed antennas - shortened verticals, vertical beams, slopers and inverted L's

- use 4:1 HCU50, 1.78:1 HDU50, etc.
- Shunt fed towers performing as verticals - use 1.56:1 HDU50
- Beverage antennas - use W2FMI-BEV

"The unun should find some important applications. Until recently, they have been literally impossible to obtain. But now they can be designed to match 50-ohm cable to loads as small as 3.125 ohms and as high as 800 ohms (both unbalanced loads). They can come in low-power and high-power versions as well as

multimatch designs. An important application for ununs is to match 50 ohm cable to the input impedance of ground-fed antennas such as verticals, slopers, inverted L's and towers performing as verticals. Since ground losses play a large role with these systems, a knowledge of the input impedance of the antenna as well as the ground loss is important in determining the optimum impedance ratio. Also, a high impedance, multimatch unun can be easily designed to match 50 ohm cable to the input impedance of Beverage antennas."

"Baluns (balanced to unbalanced transformers) and ununs (unbalanced to unbalanced transformers) belong to a class of matching devices known as transmission line transformers. They transmit the energy from input to output by a transmission line mode instead of by flux linkages as in the case of conventional transformers. When properly designed, they can have extremely high efficiencies and very broad bandwidths."

A 2:1 unun is a good choice for feeding a 1/4-wavelength "Inverted L" type antenna. For example:
One 2:1 unun they offer has 22/25-Ohm taps on one side, and a 50-Ohm tap on the other side.
This aluminum cased model (HDU100) has female SO-239 connectors at each end, too.

C&S Engineering
9229 Goldenrod Drive
Fort Wayne, IN 46835
Phone Orders: (219) 485-1458

E-mail: Chuck, N8BYI

Web: **C&S Engineering**

C&S Engineering offers several products, including their "Front End Saver (FES)," available as a kit ($22.95); pre-populated curcuit board ($29.95); or built, tested and cased ($49.95), plus shipping costs. The FES "protects your receiver when you operate on the aux. (RX) antenna. A must for 80m and 160m operators." (The FES was described in *CQ* Magazine, February 1997, p 32.)

In addition, C&S offers a "160 Meter Preamp," (also available for 80 meters). Single-band preamps are available, in kit form ($29.95); built, tested, and cased ($59.95); or for 80 *and* 160--switchable--($99.95). These preamps offer tuned inputs and outputs. "Amplify sig not noise. Over 25 dB gain, lo noise, gain control. A must for Beverage and loop antennas." (The 160-meter preamp was featured in the October 1989 issue of *Ham Radio* Magazine, in an article by Gary Nichols, KD9SV.)

Here's an ad I saw in the "Ham Ads" in the back of *QST* Magazine:

"BEVERAGE matching transformer kits, 160/80/40 DX reception.
$12 each, 3/$29. Lectrokit, 401 W. Bogart, Sandusky, OH 44870."

ARRL Books:
"Low Band DXing:"
Low band antennas (TX and RX, including Beverages), operating techniques; by ON4UN

"DXing on the Edge--The Thrill of 160 Meters:"
160-meter operating, TX and RX antennas, and more; by K1ZM (includes an audio CD, too).

Beverage antenna articles--from *QST*:
"Beverage Antennas for Amateur Communications," *QST* Magazine, January 1983, pp. 22-27. (Belrose, Litva, Moss, and Stevens)

"The Classic Beverage Antenna, Revisited," *QST* Magazine, January 1982, pp. 11-17
(H. H. Beverage and Doug DeMaw).

"The Wave Antenna for 200-Meter Reception," *QST* Magazine, November 1922, pp. 7-15
(H.H. Beverage).

K9AY receiving-loop antenna article--from *QST*:
"The K9AY Terminated Loop--A Compact, Directional Receiving Antenna"
QST Magazine, September 1997, pp. 43-46 (Gary Breed).

"EWE" receiving antenna articles--from *QST*:
"Is This EWE For You?," *QST* Magazine, February 1995, p 31 (Koontz).
"Feedback," (Re: Is This EWE For You?), *QST* Magazine, April 1995, p 75 (Koontz).

"More EWE's For You" *QST* Magazine, January 1996, p 32 (Koontz).

Further reading:

"The Beverage Antenna Handbook"
Victor Misek, W1WCR
142 Wason Road

Hudson, NH 03051

"Beverage and Longwire Theory"
National Radio Club
P.O. Box 164
Mannsville, NY 13661

"The Beverage Antenna"
Popular Electronics magazine
January 1998 issue; pages 40 to 46
An article by:

Joseph J. Carr, K4IPV

Receive-antenna-related links:

"Who's Who on the TopBand" by VK2ICV:
Lists the top-gun DXers on 160 meters, by number of countries worked on our "TopBand."

"Terminated Shortwave Antenna" by Cyril Dufault, VA2CJD:
" -- the "baby" [B]everage antenna for everybody --"
(It's a terminated loop, actually.)

"T2FD -- The Forgotten Antenna" About the terminated, tilted, folded dipole,
this article appears on the web page of the "Nordic Shortwave Center:"

"An early discussion of the T2FD appeared in the June 1949 issue of *QST*. The author of that article followed up on the T2FD in the November 1951 issue of *QST*, and in the February 1953 issue of *QST*. A more recent article on the T2FD appeared in the May 1984 issue of *73 Magazine*."

You'll also find links to how to construct a T2FD, using a 4:1 balun, 390-Ohm resistor, and 75-Ohm RG-59 type coaxial cable; And, a comparison of an 60 meter band T2FD (for SWLs), to a 500-foot-long longwire-type receiving antenna, and to a 50-foot-long random-wire antenna.

Page author: Glenn Swanson, KB1GW

kb1gw@snet.net

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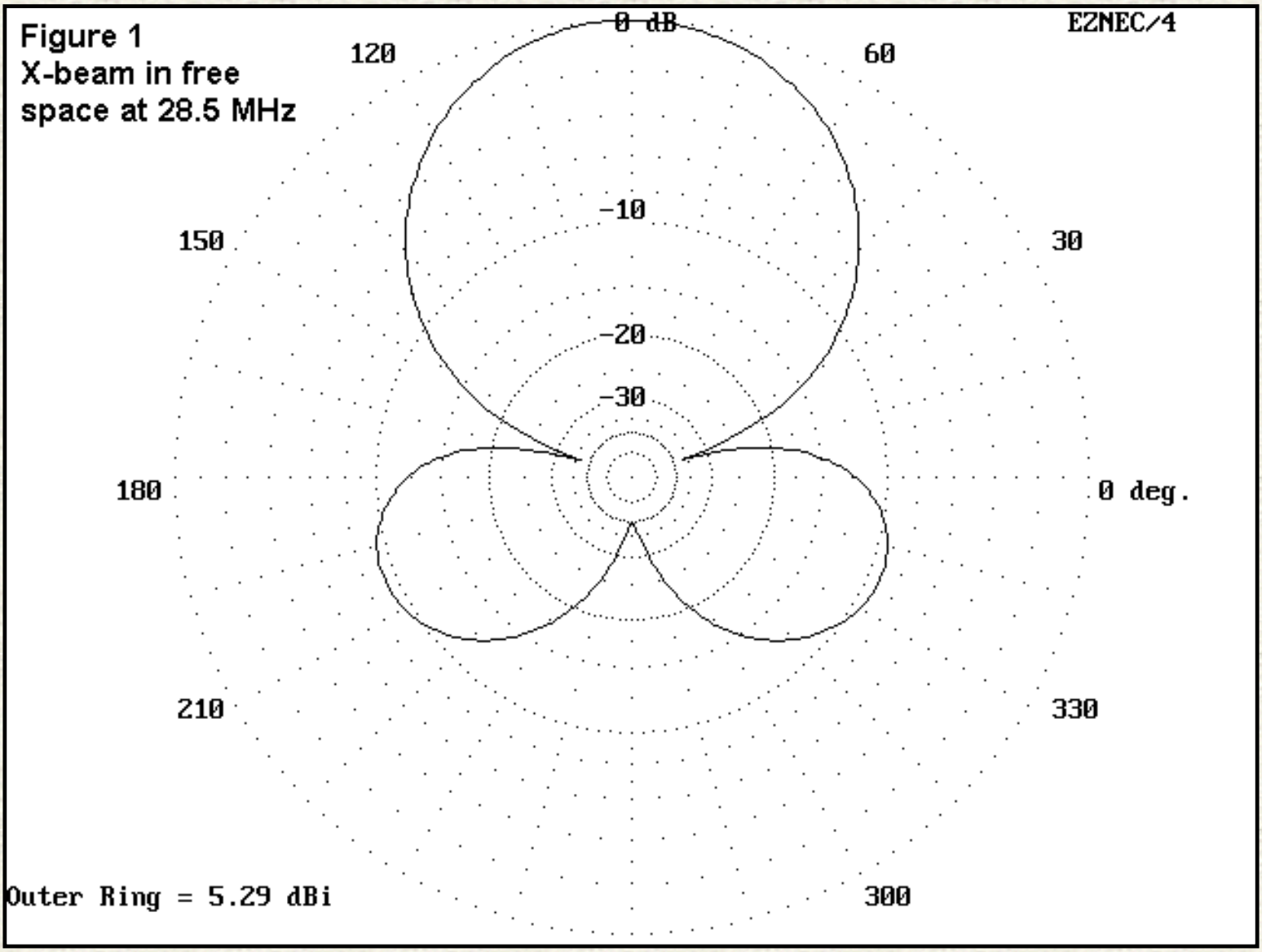
The Birdcage Antenna

L. B. Cebik, W4RNL

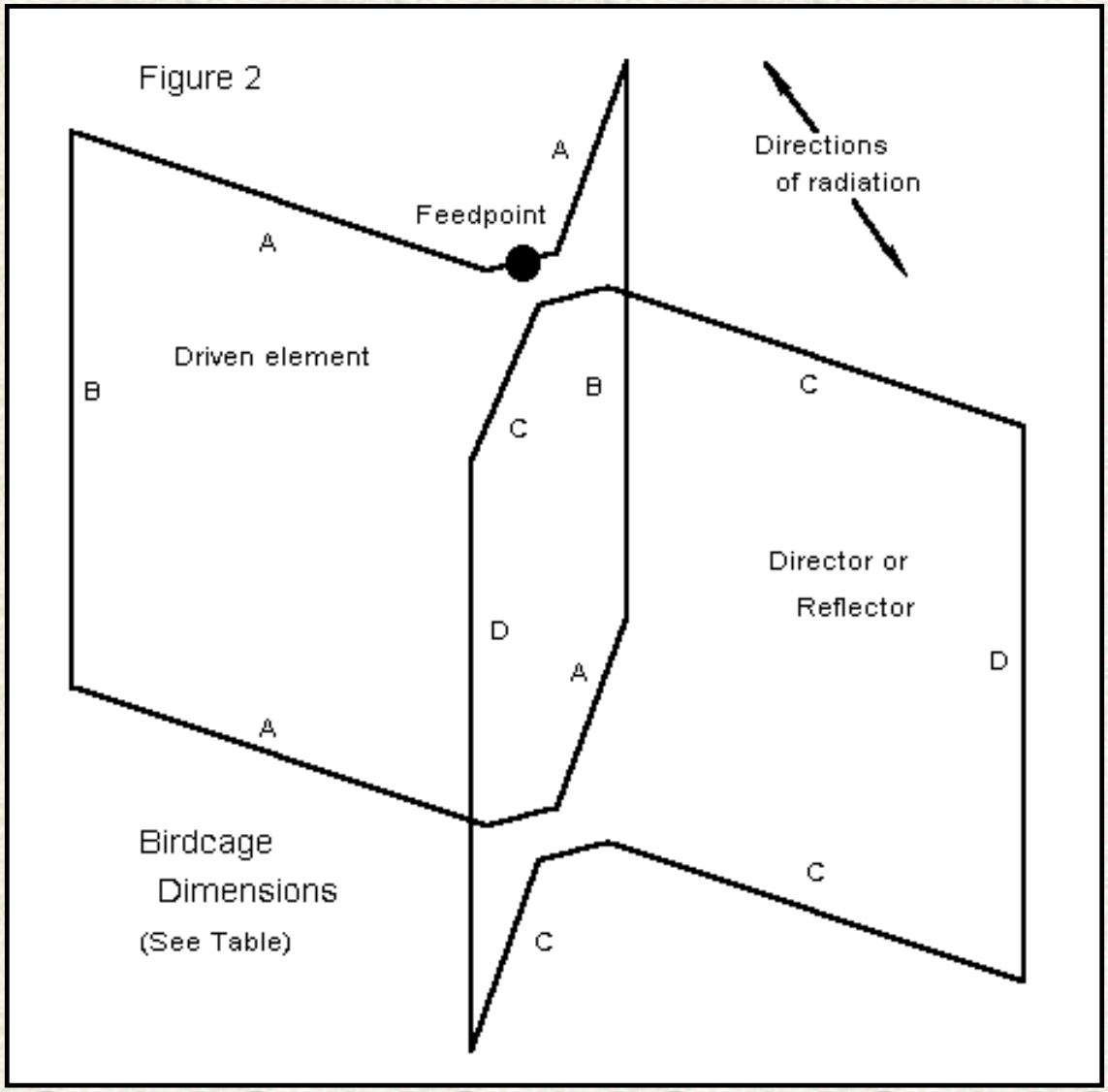
The folded X-beam derives its existence from the G4ZU Birdcage antenna. W2EEY viewed the X-beam as half a Birdcage.(1) Although this is historical fact, the logic of antennas goes the other way round: the Birdcage is a double folded X-beam. For reference, Figure 1 provides a free space pattern of an X-beam tuned for maximum front-to-back ratio.



We may legitimately ask, how much better does the whole Birdcage work compared to an X-beam? When Dick Bird,

G4ZU, published his work in 1960, he made some fairly extravagant claims for the antenna, claims that led him to patent it.(2) At this distance and with the calculating power made available by NEC and its offspring, we may decide that the claims for his antenna (and those of many hams who built one) reflect more accurately the low state of home-brew Yagi and quad construction of the time than they do contemporary comparative performance. Models of the Birdcage show little advantage over the X-beam and exaggerate some of its limitations.

Bird graphically demonstrated what he took to be the principles of the Birdcage by beginning with the bidirectional pattern of the dipole. Folded into a Vee, the antenna would show, according to his sketches, a strong lobe out the open end and a weak lobe off the apex. Add a Vee reflector (pointed the other way to make an X), double the assembly a quarter wave above or below, bend the X-arms vertically to meet each other as phasing lines, and the Birdcage is born. Figure 2 shows the outline of the Birdcage, with identification of each relevant dimension.



Unfortunately, the seeming flaw in the reasoning begins with the Vee itself. Short Vees do not cancel the radiation off the backsides of the elements (the outsides of the V). Thus are born the uneliminable quartering lobes of the X-beam and even larger lobes for the Birdcage.

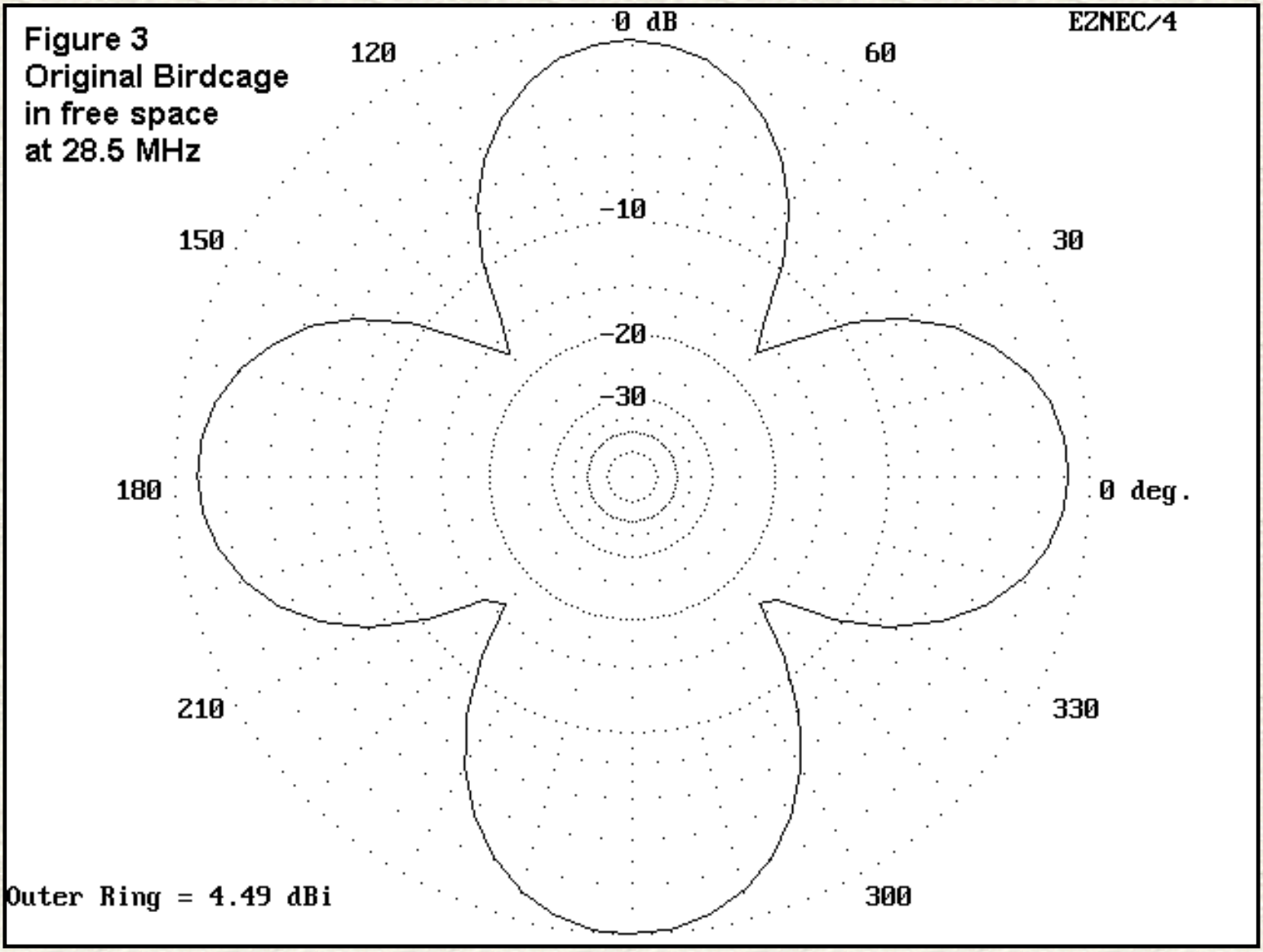
Birdcages come in two forms: with a reflector and with a director. Table 1 shows 10-meter models of each beginning with a self-resonant Birdcage with equal-size elements derived from data given in 1960's literature.(3)

Table 1. Modeled Dimensions of the Birdcage Antenna

Dimensions	DE Arm (A) L (ft)	Parasitic (C) L (ft)	Vertical (B and D) L (ft)
Basic Birdcage	5.05	5.05	8.666
DE and Refl Resonated	4.95	5.26	8.666
DE and Dir Resonated	5.16	4.91	8.666

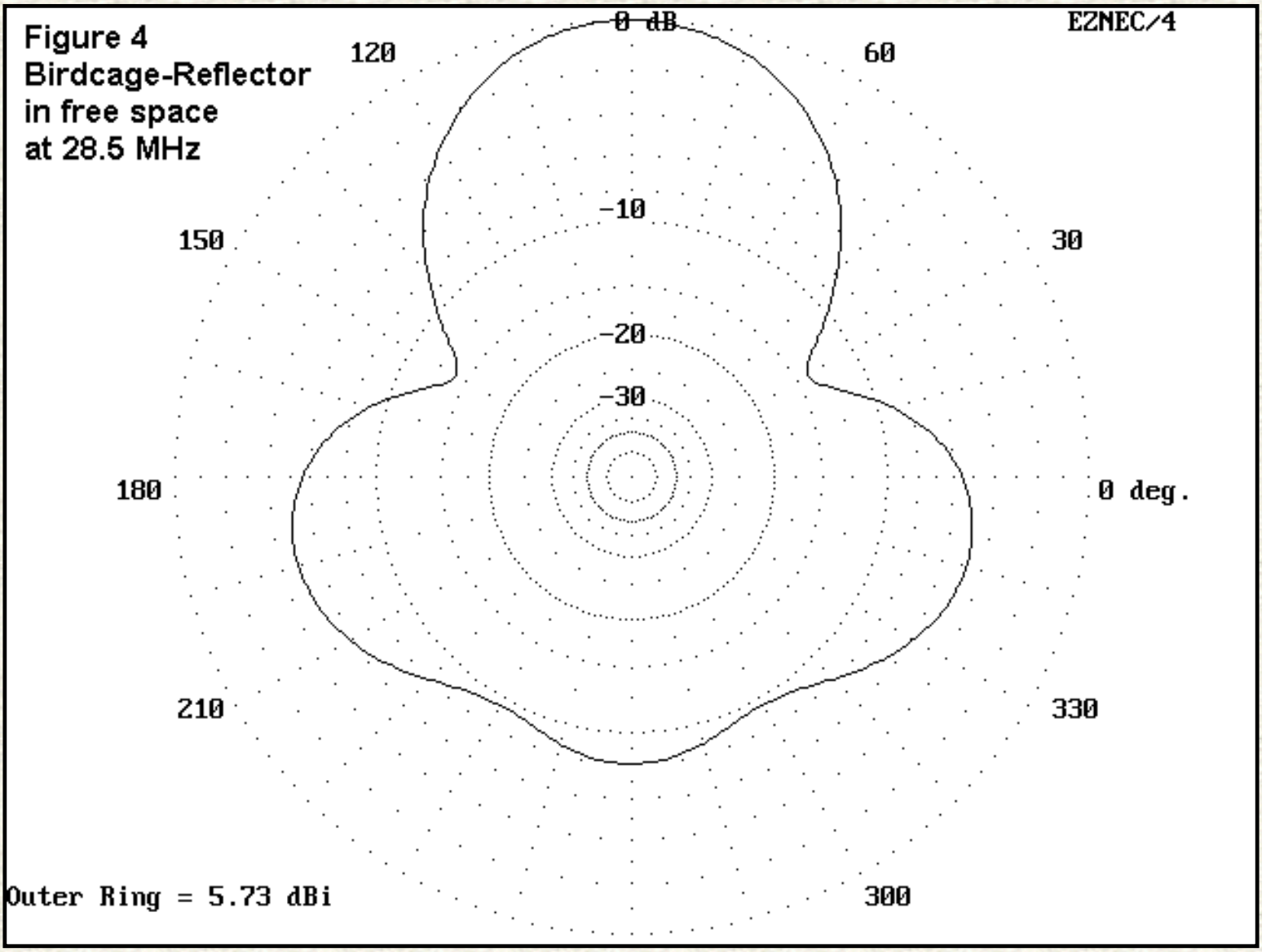
Note: all models used #14 copper wire in free space at 28.5 MHz on NEC-4.

Figure 3 shows the performance in free space of the basic Birdcage--a nice 4-leaf clover.

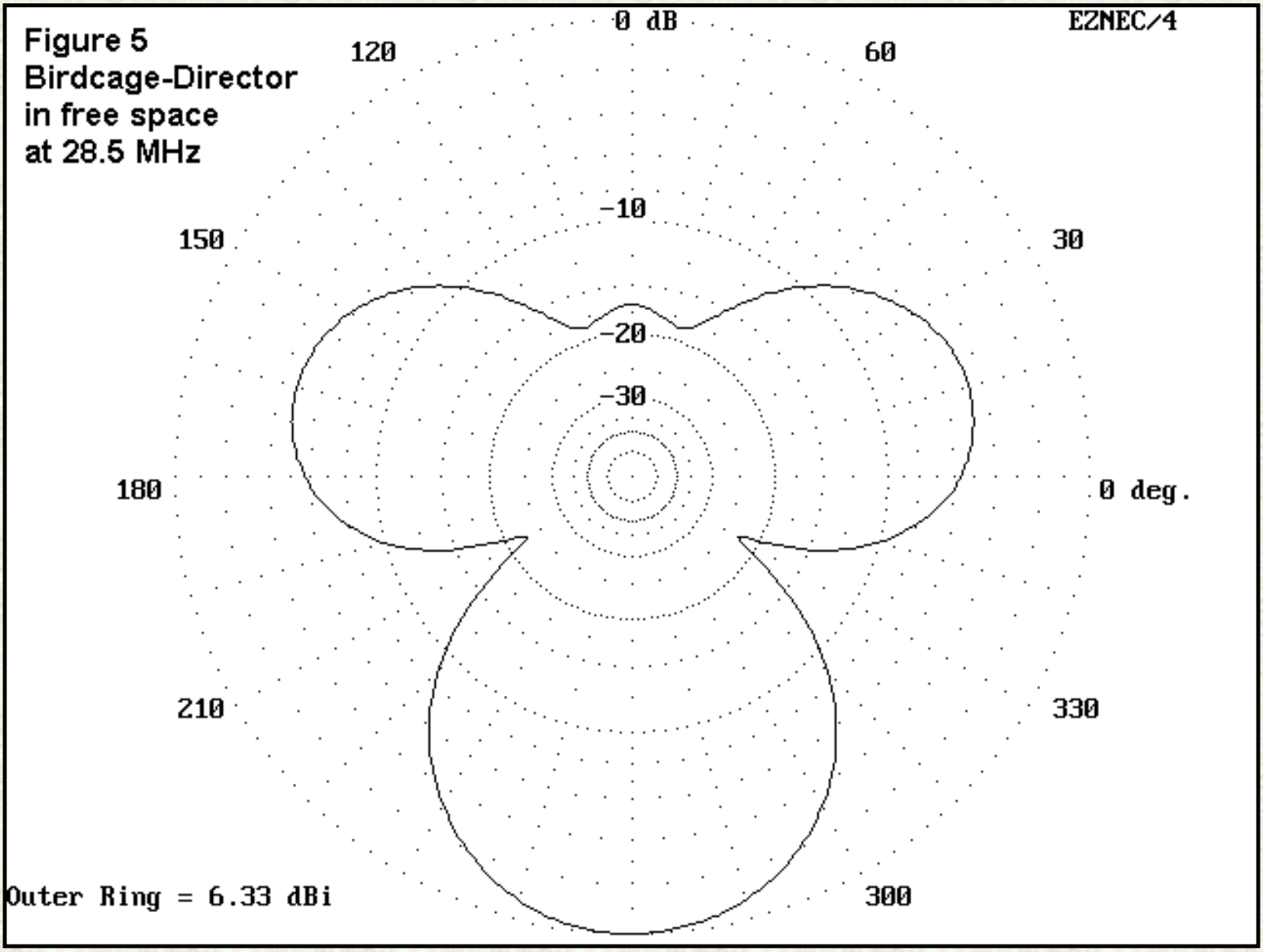


G4ZU favored the reflector version, which is perhaps the worse performer of the two possible versions. With a

forward gain in free space of under 6 dBi and a front-to-back ratio of about 8 dB, the beam is not the equal of a 2-element Yagi or even of the X-beam. Figure 4 shows its performance in free space.



The director version that parallels the development of the X-beam provides more gain and front-to-back ratio: 6.3 dBi and 17 dB, respectively. Figure 5 shows its performance in free space. The direction is reversed, since I simply re-dimensioned the reflector to form a director.



As shown in Table 2, the figures for the directed Birdcage surpass both the X-beam and the 2-element Yagi by a small bit, while the reflected Birdcage leaves much to be desired.

Table 2. A Comparison of Antenna Performance

Antenna Type	TO Angle (degrees)	Gain (dBi)	Front-to-Back (dB)	Feedpoint R ñ jX (Ohms)
A. Free Space				
Birdcage: DE + Ref	--	5.8	8.0	34 - 0
Birdcage: DR + Dir	--	6.3	16.9	40 + 4
Folded X-Beam	--	5.3	39.1	37 - 3
B. Height = 5/8 wl				
Birdcage: DE + Ref	22	10.5	8.0	31 - 3
Birdcage: DR + Dir	22	11.0	17.1	40 + 3
Folded X-Beam	23	10.5	21.8	35 + 0
2-El Yagi	23	11.3	10.9	34 - 5
2-El Quad	22	11.7	18.3	97 - 3
C. Height = 1 wl				
Birdcage: DE + Ref	13	11.3	8.6	32 - 2

Birdcage: DR + Dir	**14**	**11.8**	**17.0**	**39 + 2**
Folded X-Beam	**14**	**10.8**	**31.7**	**33 - 1**
2-El Yagi	**14**	**11.7**	**13.5**	**35 + 0**
2-El Quad	**13**	**12.4**	**17.9**	**92 + 8**

Note: TO angle=the angle of maximum radiation at which azimuth readings of calculated gain and front-to-back ratio were obtained. All calculations via NEC-4.

The pattern of the Birdcage in either version does not meet today's standards of front-to-side ratio or front-to-rear ratio. The quartering rear lobes of the X-beam become virtual side lobes of the Birdcage and are only down from the main lobe by some 4 to 5 dB in the either version. There is little QRM-rejection potential, especially in the reflector version. Over real ground, the pattern devolves into a mound of whip cream. However, over real ground, the director version tends to push its side lobes to the rear, and the pattern begins to resemble that of the folded X-beam.

Like the folded X-beam, the Birdcage retains its most optimal pattern only over a small portion of the band of design. The reflector version rapidly grows a tail below design frequency, while the director version grows its tail above design frequency. The SWR bandwidth is also narrow, less than 0.5 MHz at 10 meters for a 2:1 SWR. However, the complex structure of the Birdcage is less amenable to a tunable parasitic element than is the simpler X-beam.

All in all, the additional complexities of the Birdcage are not warranted by a significant increase in performance over the X-beam. Gain aside, the side and rear performance of the antenna falls far short of other compact designs. Consequently, while that antenna has a great name and an appealing look, performance models do not recommend the investment of time and resources in this antenna. In some ways, it is a shame to discover this: G4ZU contributed extensively to amateur practice in the field of antennas. Although the Birdcage antenna has had its day, Dick's other work--and his example--will long endure.

Notes

(1) John Schultz, W2EEY, "The G4ZU X Beam for 20," CQ (June, 1965), pp. 26- 28, 102.

(2) Dick Bird, G4ZU, "The G4ZU 'Bird Cage' Aerial," CQ (April, 1960), pp. 40-42, 117.

(3) See George Cousins, VE1TG, "The Tri-Band Birdcage," CQ (July, 1963), pp. 30-33. Another brief reference to the antenna occurs in Orr and Cowan, Cubical Quad Antennas, 3rd Ed. (Lakewood, NJ: Radio Amateur Callbook, 1993), p. 102, where it is shown with a reflector and referred to as a variation of the quad.

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SPACING. - The lower relative efficiency of collinear arrays of many elements, compared with other multi-element arrays, relates directly to spacing and mutual impedance effects. Mutual impedance is an important factor to be considered when any two elements are parallel and are spaced so that considerable coupling is between them. There is very little mutual impedance between collinear sections. Where impedance does exist, it is caused by the coupling between the ends of adjacent elements. Placing the ends of elements close together is frequently necessary because of construction problems, especially where long lengths of wire are involved.

The effects of spacing and the advantages of proper spacing can be demonstrated by some practical examples. A collinear array consisting of two half-wave elements with 1/2-wavelength spacing between centers has a gain of 1.8 dB. If the ends of these same dipoles are separated so that the distance from center to center is 3/4 wavelengths and they are driven from the same source, the gain increases to approximately 2.9 dB.

A three-dipole array with negligible spacing between elements gives a gain of 3.3 dB. In other words, when two elements are used with wider spacing, the gain obtained is approximately equal to the gain obtainable from three elements with close spacing. The spacing of this array permits simpler construction, since only two dipoles are used. It also allows the antenna to occupy less space. Construction problems usually dictate small-array spacing.

Broadside Arrays

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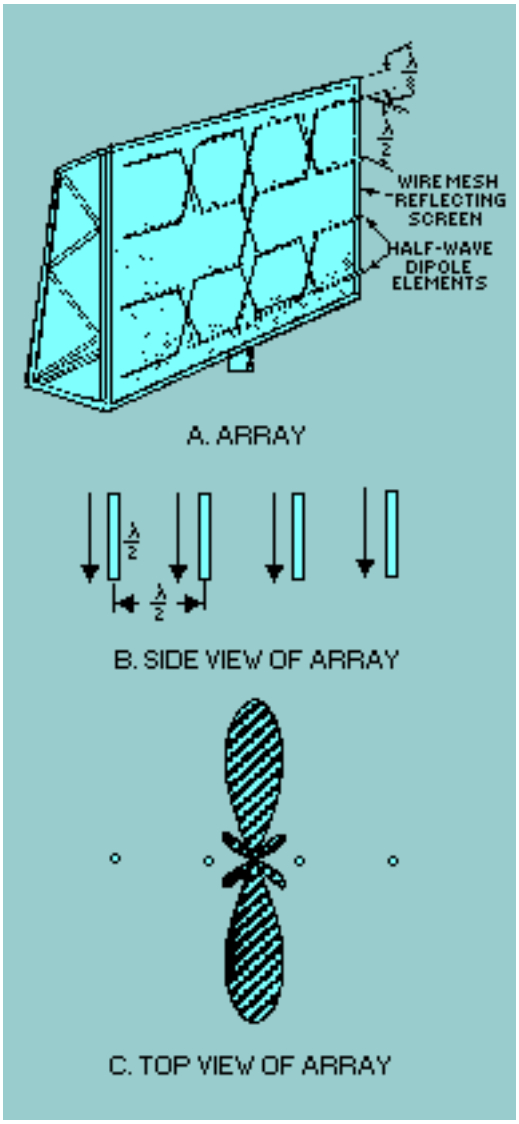
A broadside array is shown in figure 4-26, view A. Physically, it looks somewhat like a ladder. When the array and the elements in it are polarized horizontally, it looks like an upright ladder. When the array is polarized vertically, it looks like a ladder lying on one side (view B). View C is an illustration of the radiation pattern of a broadside array. Horizontally polarized arrays using more than two elements are not common. This is because the requirement that the bottom of the array be a significant distance above the earth presents construction problems. Compared with collinear arrays, broadside arrays tune sharply, but lose efficiency rapidly when not operated on the frequencies for which they are designed.

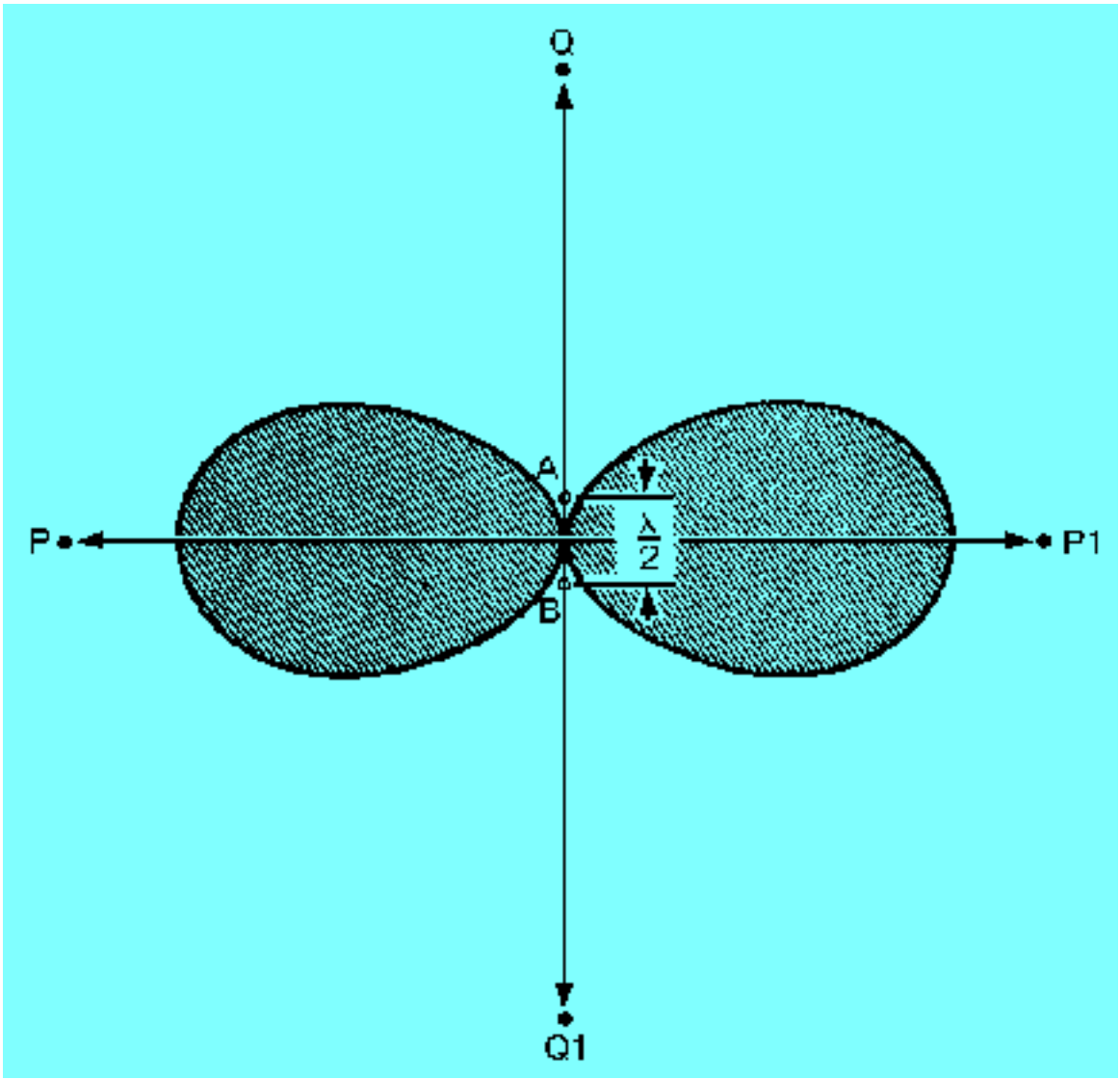
Figure 4-26. - Typical broadside array.



RADIATION PATTERN. - Figure 4-27 shows an end view of two parallel half-wave antennas (A and B) operating in the same phase and located 1/2 wavelength apart. At a point (P) far removed from the antennas, the antennas

appear as a single point. Energy radiating toward P from antenna A starts out in phase with the energy radiating from antenna B in the same direction. Propagation from each antenna travels over the same distance to point P, arriving there in phase. The antennas reinforce each other in this direction, making a strong signal available at P. Field strength measured at P is greater than it would be if the total power supplied to both antennas had been fed to a single dipole. Radiation toward point P1 is built up in the same manner.

Figure 4-27. - Parallel elements in phase.



Next consider a wavefront traveling toward point Q from antenna B. By the time it reaches antenna A, 1/2 wavelength away, 1/2 cycle has elapsed. Therefore energy from antenna B meets the energy from antenna A 180 degrees out of phase. As a result, the energy moving toward point Q from the two sources cancels. In a like manner, radiation from antenna A traveling toward point Q1 meets and cancels the radiation in the same direction from antenna B. As a result, little propagation takes place in either direction along

the QQ1 axis. Most of the energy is concentrated in both directions along the PP1 axis. When both antenna elements are fed from the same source, the result is the basic broadside array.

When more than two elements are used in a broadside arrangement, they are all parallel and in the same plane, as shown in figure 4-26, view B. Current phase, indicated by the arrows, must be the same for all elements. The radiation pattern shown in figure 4-26, view C, is always bidirectional. This pattern is sharper than the one shown in figure 4-27 because of the additional two elements. Directivity and gain depend on the number of elements and the spacing between them.

GAIN AND DIRECTIVITY. - The physical disposition of dipoles operated broadside to each other allows for much greater coupling between them than can occur between collinear elements. Moving the parallel antenna elements closer together or farther apart affects the actual impedance of the entire array and the overall radiation resistance as well. As the spacing between broadside elements increases, the effect on the radiation pattern is a sharpening of the major lobes. When the array consists of only two dipoles spaced exactly 1/2 wavelength apart, no minor lobes are generated at all. Increasing the distance between the elements beyond that point, however, tends to throw off the phase relationship between the original current in one element and the current induced in it by the other element. The result is that, although the major lobes are sharpened, minor lobes are introduced, even with two elements. These, however, are not large enough to be of concern.

If you add the same number of elements to both a broadside array and a collinear array, the gain of the broadside array will be greater. Reduced radiation resistance resulting from the efficient coupling between dipoles accounts for most of this gain. However, certain practical factors limit the number of elements that may be used. The construction problem increases with the number of elements, especially when they are polarized horizontally.

Q.32 What is the primary cause of broadside arrays losing efficiency when not operating at their designed frequency? Answer
Q.33 When more than two elements are used in a broadside array, how are the elements arranged? Answer
Q.34 As the spacing between elements in a broadside array increases, what is the effect on the major lobes? Answer

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RADIO HABANA CUBA

The Broomstick Special: Part 1

A compact, easy-to-build shortwave antenna

BY ARNIE CORO
Host of "Dxers Unlimited"

SEND YOUR COMMENTS, QUESTIONS AND IDEAS DIRECTLY TO ME AT:
arnie@radiohc.org

Yes, this is a "helically wound" wire antenna. It can be built in a few hours... it will take longer if you really want to make it look pretty... UGLY versions can be assembled in minutes. It MAY be used without an antenna tuner, BUT... it works best when you do USE a TUNER.

Here are the easy-to-follow, step-by-step building instructions for Arnie Coro's "Broomstick Special:"

1. Obtain a nice broomstick.... you can use a "classic" wooden dowel broomstick, or substitute heavy walled PVC plumbers plastic tubing

2. Diameter of broomstick is not critical; anything from about 1.5 centimeters or better will work (this means that PVC tubing of about 19mm or 3/4 inch is ideal)

3. Prepare a base to hold the broomstick or PVC pipe vertical... Use a wide base, with enough counterweight attached to keep the broomstick vertical (I use mine next to the bedside radio, have convinced the wife that it is "modern art"

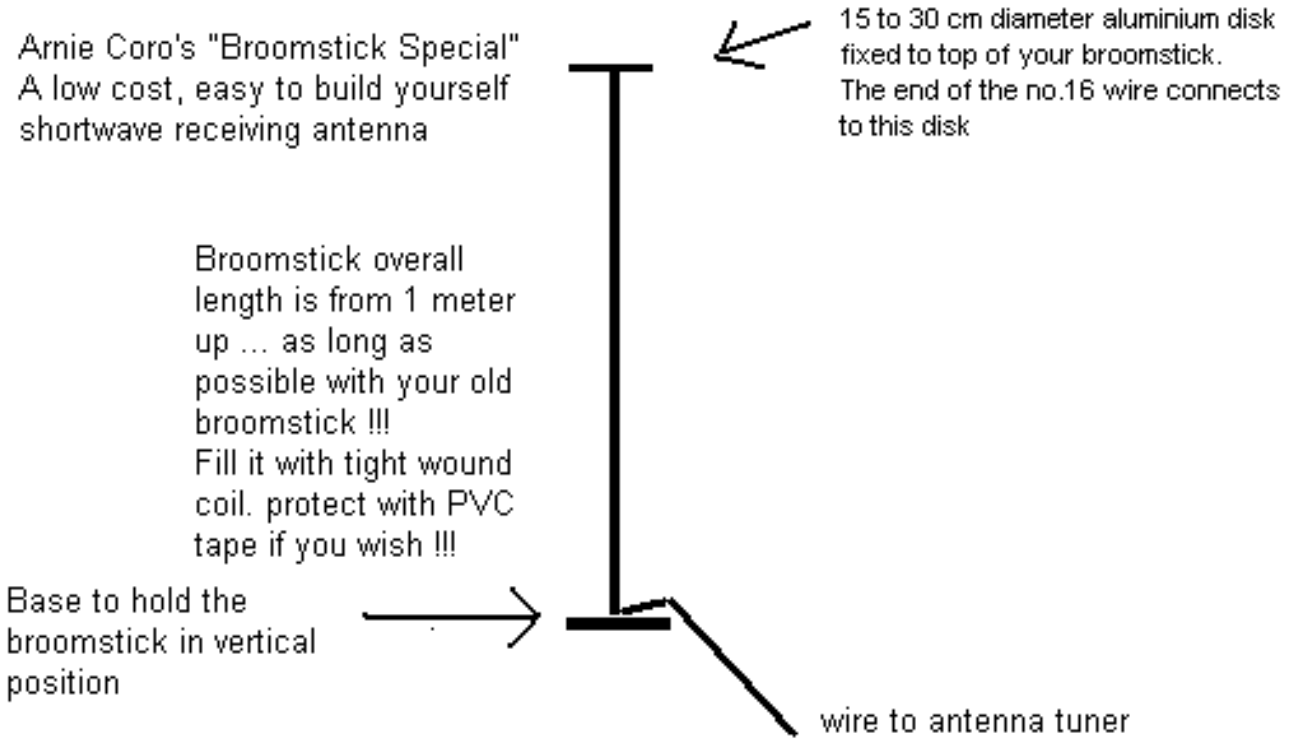
4. Obtain an aluminium disk of no less than 15 cm (6 inches) diameter. I prefer using a disk of around 30 centimeters (12 inches) but this is not critical. DO USE THE DISK... as it provides a capacity hat termination and helps reduce NOISE PICK UP

5. Obtain enough No.16 PVC plastic covered household wire; this is the ideal choice, but if you can't find it, then you may use No.16 or No.18 enamel covered copper wire (the one used for winding motors and transformers). If you can't find No.16 PVC covered wire, then your second best choice is No.18 "speaker wire"

6. Connect one end of the wire to the aluminium disk, and start winding a uniform coil using the "broostick" as the coil form. YOU WANT A NEAT JOB! Turns should fit tightly one next to the other... the "broomstick" will be filled with the wire forming the coil... When you arrive at the bottom end, make a termination> I use a long bolt with

nuts and washers to which I tie the end of the wire, and another wire that goes to the antenna tuner. This wire that goes to the tuner can be from 1 meter to 3 meters long (from 3 to 10 feet) but DO KEEP IT AS SHORT AS POSSIBLE.

7. After the antenna is built, you may want to protect it with tightly wound PVC plastic tape over the wire. For EXTRA protection, you can paint the whole antenna with several coats of SPRAY ENAMEL...



8. The antenna works best near a WINDOW!!! Or better yet, you can install it in your balcony or garden... BUT DO KEEP THE CONNECTIONS TO THE TUNER SHORT

9. The antenna is RESONATED with your antenna tuner.... YOU MAY USE IT WITHOUT A TUNER but results are not going to be as good as when the antenna is connected to the receiver via a well-designed antenna tuner

10. YES... YOU MAY USE IT FOR TRANSMITTING.... BUT... according to recent medical research information, keep it as far away as possible from your body!!!

11. The helically wound "broomstick" is a lot of fun to experiment with... It works best when you provide a ground connection to the antenna tuner - receiver combination. One way of providing an "artificial ground" is to connect a length of wire of no less than 5 meters as a "counterpoise," that meaning that you can let the wire hang around the room's floor or garden. Using the "broomstick" with resonant radials turns it into an excellent amateur radio antenna for a specific band... For example with 4 radials cut for

the 15 meter amateur band and a 4 feet high broomstick, (about 1.5 meters)

I can work a lot of stations on 21 mHz, something I do often to demonstrate to visitors what can be achieved with simple homebrew antennas, even when you don't have a lot of space.

12. QUESTIONS.... to: **arnie@radiohc.org**

13. FOR THE TECHNICALLY MINDED... ***Click here for Part 2 of The Broomstick Antenna!***

Havana, Cuba
6 February, 1998

Send e-mail to Arnie Coro

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Welcome to the W3FF Home Page

Home of the homebrew "Buddipole" - A Portable Dipole Antenna Design

Sorry, your browser does not support Java

This site is for those hams who are seeking a high performance homebrew antenna for hf portable use. As you will see, I am particularly interested in pedestrian mobile operating. I started out in early 2000 with a Gap backpack, a TS-50, and a 7.5 amp-hour, six pound battery. Because I wanted to take my radio with me and communicate on HF, I designed a portable dipole antenna out of easily-found and low-cost materials. The Buddipole is the result of those efforts. This is a lightweight, modular, and rotatable dipole (a great benefit, by the way) and is very efficient on the higher bands. See the links below for information on how to build this popular antenna.

The original antenna design is still great. But there were problems with the PVC plastic pipe. It sags. The connectors become loose. The Radio Shack whips are flimsy, and they bend easily. So, the homebrew Buddipole project took on another level of sophistication. The result is a commercial antenna with the same name. You can find it here: www.buddipole.com.

New materials, like epoxy composite and polypropylene, and 6061 aircraft aluminum, and Delrin. One set of higher-Q coils to work all the bands of the original version, plus 30m and 40m. Parts are custom-machined and we had new custom stainless steel whips manufactured to our specifications. We use stretch Velcro as a strain relief for the new coax rf bead assembly, and the assembly is encased in an adhesive shrinkwrap. Even the brass studs are drilled out to lighten them without sacrificing durability. We have a Delrin adaptor that takes painter's pole threads (Acme) to pipe threads that fit right into the new poly T's on the commercial product.

And now, we have the B.U.G. -- A Buddipole User's Group. You can sign up for this Yahoo Group on this page. A truly great resource to share many of the ideas, experiences, and suggestions for both the homebrew and commercial Buddipole.

Build the homebrew version on this site. Modify it. Tell us what you have worked with it on the air. We are always thrilled to hear from folks who build the antenna and who are active on the air with it! And consider upgrading to the new version, with the Rotatable Arms, that gives you a choice of six different configurations to play with.

Have fun and 73,

W3FF - Budd

Join the Buddipole User's Group!

Powered by groups.yahoo.com

This is a new internet based discussion group that began on July 19, 2002. It is an excellent resource for Buddipole users to share their thoughts, ideas, experiences, photos, and anything else that is Buddipole related. Drop by to see what everyone is talking about!

The W3FF Buddipole Antenna

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Hear it

See it

Build the Buddipole

Buddipole FAQ Page

Dayton 2002 Recap

Join the **W3FF Buddipole** email list

Email:

This is our email newsletter that we send to subscribers with tips, ideas, and suggestions for using the Buddipole. In addition, we will announce new products via this newsletter.

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The HFpack - The HF Portable Group
S.C.A.R.S. HOME PAGE (Local Club)
QRZ The Live-Wire Group - Ham experimental antennas, portable operations

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Five RG59 Coax Bundle

Miniature Coaxial Cable
Miniature Coaxial Cable
Miniature 3 & 5 Coax Bundle

Cable Bundles
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Five RG-59/U
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Two RG-6/U Quad + Two Cat 5e + Two Fiber
Three or Five Miniature Coax

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Earn TechCash	**BE-1189A** COAX RG-6/U Quad Shield NEC CATV CL2		36 lbs/spool/1000ft **$129.00 spool/1000ft**		Enter Qty spool/1000ft		IN STOCK
Earn TechCash	**BE-1189AW** COAX RG-6/U Quad - White NEC CATV CL2	List 0.32 foot	**$0.26 foot**		Enter Qty foot		LOW STOCK
Earn TechCash	**BE-1189AW** COAX RG-6/U Quad - White NEC CATV CL2		36 lbs/spool/1000 **$129.00 spool/1000**		Enter Qty spool/1000		LOW STOCK

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- **Note:** shipping charges for bulk wire and cable are actual cost.

Earn TechCash	**BE-9116** COAX RG-6/U Dual Shield NEC CATV CL2 Black	List 0.22 foot	**$0.18 foot**		Enter Qty foot	IN STOCK
Earn TechCash	**BE-9116** COAX RG-6/U Dual Shield NEC CATV CL2 Black		30 lbs/spool/1000ft **$87.90 spool/1000ft**		Enter Qty spool/1000ft	IN STOCK

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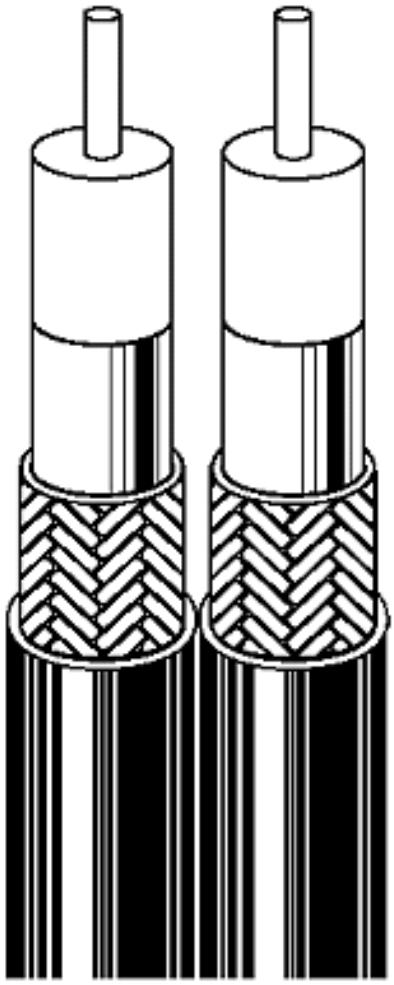
RG-6/U Dual Shield Siamese Coaxial Cable



- Use RG-6/U Siamese Dual Shield for general purpose video cable for distribution of broadcast and cable TV throughout your home whenever you need to run two cables at a time.
- Perfect for dual LNB satellite dish installations!
- Available by the foot and in 1000' spool.
- Full Specifications (GIF)
- **Note:** shipping charges for bulk wire and cable are actual cost.

Earn TechCash	**BE-9077** RG6,Siamese,Dual Shield NEC CATV CL2	List 0.54 foot	**$0.45 foot**		Enter Qty foot	IN STOCK
Earn TechCash	**BE-9077** RG6,Siamese,Dual Shield NEC CATV CL2		66 lbs/spool/1000ft **$219.00 spool/1000ft**		Enter Qty spool/1000ft	IN STOCK

MONSTER CABLE

MVQ

RG-6/U Quad-Shield Coaxial Cable

- Designed for high quality transmission with minimal interference.
- Quad shield minimizes both air interference and low level AC.
- Thin copper strands allow for one of the most flexible Quad Shielded coax cables on the market.
- Powder coated outer coat for smooth surface and easy runs.
- Available by the foot and in a 500' spool.
- **Note:** shipping charges for bulk wire and cable are actual cost.

Earn TechCash	**MC-MVQ** RG-6 Quad Shield	List 0.80 foot	**$0.77 foot**		Enter Qty foot	TEMP OUT OF STOCK
Earn TechCash	**MC-MVQ** RG-6 Quad Shield		30 lbs/spool/500ft **$219.95 spool/500ft**		Enter Qty spool/500ft	TEMP OUT OF STOCK

MONSTER CABLE

ISATR

RG-6/U ISAT Reference Coaxial Cable

- Satellite tuned.
- UV resistant.
- Direct burial rated.
- Available by the foot and in 1000' spool.
- **Note:** shipping charges for bulk wire and cable are actual cost.

Earn TechCash	**MC-ISATR** Monster Satellite ISAT REF 75 ohm High Resolution Coax	List 1.30 foot	**$1.22 foot**		Enter Qty foot	IN STOCK
Earn TechCash	**MC-ISATR** Monster Satellite ISAT REF 75 ohm High Resolution Coax		20 lbs/spool/500ft **$344.95 spool/500ft**		Enter Qty spool/500ft	IN STOCK

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RG-59/U Coaxial Cable

98132

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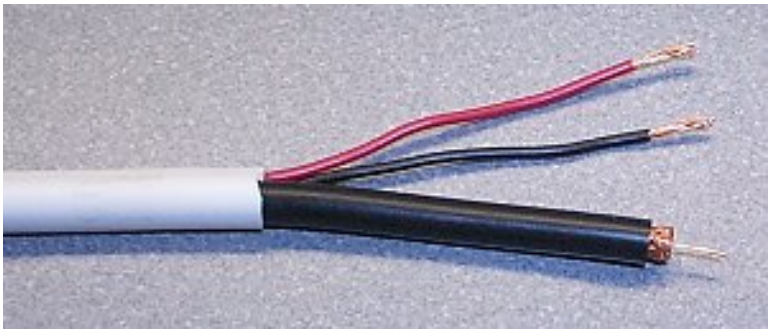
1 RG/59U + 2 18AWG Power Wires

- One RG-59 single-shield 75ohm coaxial cable.
- Two 18 gauge stranded wires.
- In a PVC Class 2 rated jacket.
- Perfect for wiring video cameras: Use the RG59 for the video feed and the red and black wires for power to the camera.
- Available by the foot and in 500' spool.
- **Note:** shipping charges for bulk wire and cable are actual cost.



Earn TechCash	**AW-98132** RG59 and 18/2 CCTV Cable	List 0.72 foot	**$0.69 foot**		Enter Qty foot	IN STOCK
Earn TechCash	**AW-98132** RG59 and 18/2 CCTV Cable		37 lbs/spool/500ft **$189.95 spool/500ft**		Enter Qty spool/500ft	IN STOCK

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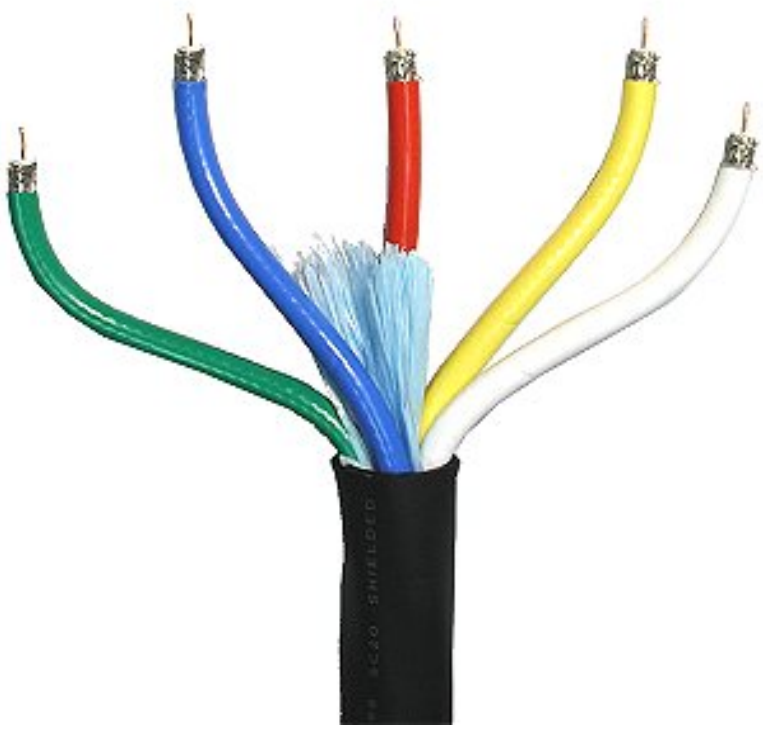
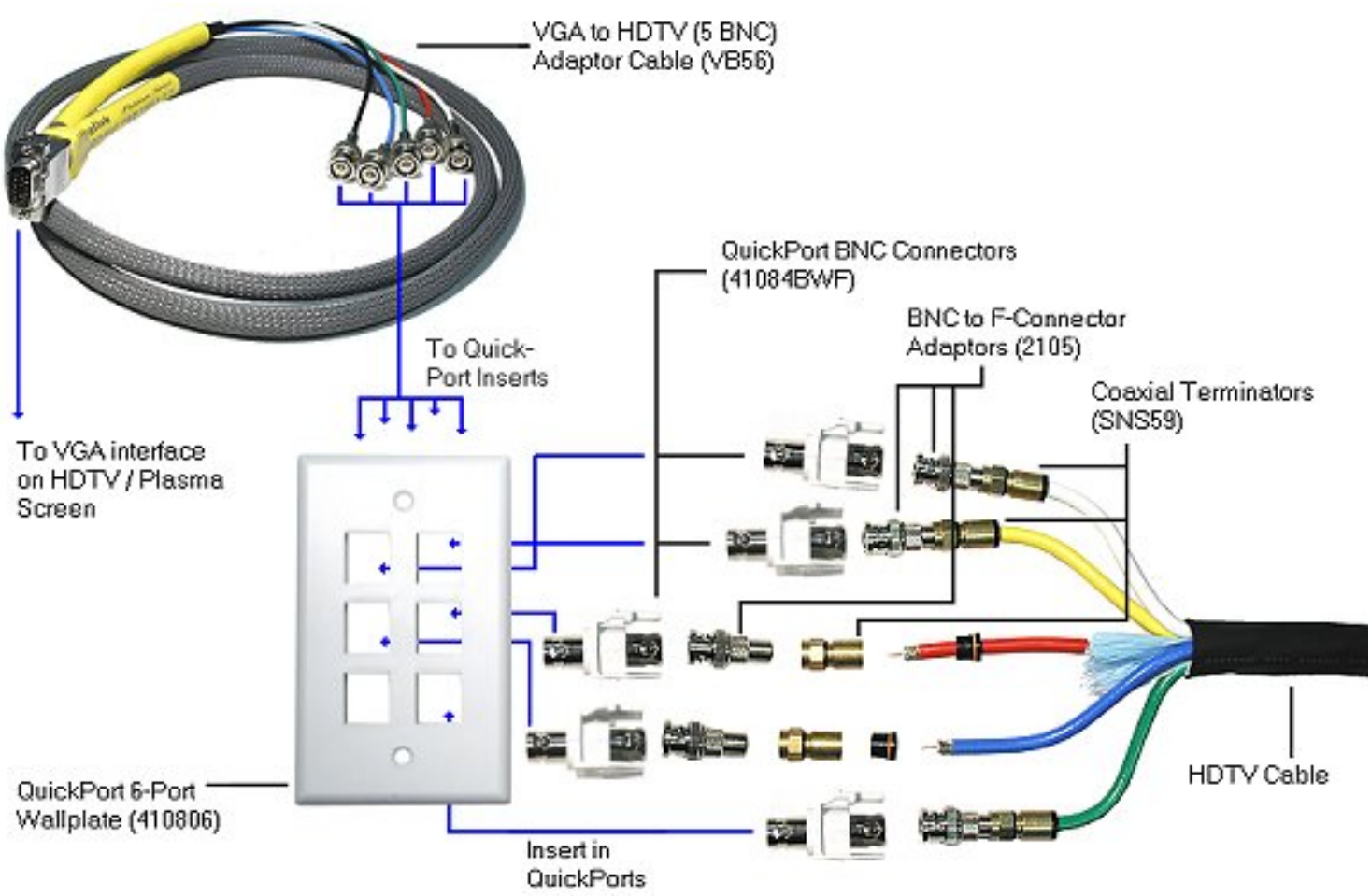
7796A

Five RG-59/U Cable Bundle (HDTV Bulk Cable)



- Belden Quality
- 5 RG59 Cables
- Color coded for easy Identification
- Low-Friction Jacket and filler for consistant shape.
- Easy runs.

This cable can be terminated to BNC male connectors in a two step process. First, strip the coax ends using an IT1000 Stipping/Compression Tool (or similar). Use the SNS59 connectors (along with the IT1000 tool) to terminate the ends. After the ends have been properly terminated, attach the BNC to F-Connector Adapters to each end. See Online Tutorial on terminating coaxial cables using the Snap N' Seal connectors.



You can use the Leviton QuickPort wallplate and the Barrel BNC QuickPort Inserts to professionally the in-wall cable and adaptors.

If your screen has only a 15-pin VGA interface, you need to purchase a UXGA to HDTV (5 BNC) Cable Interconnect. Otherwise, you can connect using BNC Patch Cables.

Earn TechCash	**BE-7796A** Bundled RG59 x 5 HDTV Cable	List 4.00 foot	0.3 lbs/foot **$3.35 foot**		Enter Qty foot	IN STOCK
Earn TechCash	**BE-7796A** Bundled RG59 x 5 HDTV Cable		150 lbs/spool/500ft **$1415.00 spool/500ft**		Enter Qty spool/500ft	IN STOCK

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Miniature Coaxial Cable

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Brilliance Miniature Dual-Shield Coaxial Cable

Only .16" (< 3/16") in diameter!

- 75 ohm coaxial cable
- 23 AWG solid bare copper center conductor
- Dual shield
- Black suitable for indoor and outdoor aerial applications. All other colors suitable for indoor use only.
- Full Specifications (GIF)
- Connectors available: "F", RCA, & BNC
- Use our 200004 stripper and 300007 crimper.

An excellent coaxial cable for running baseband video (one cable required), s-video (two cables required), component video (three cables required), or UXGA video (five cables required). Also works great for line-level audio and subwoofer audio. The small size of this cable makes it easy to route and hide. The low loss makes long runs >300ft possible. **Not suitable for raw satellite, UHF/VHF, or CATV signals.**

Earn TechCash	**BE-1855AE** Miniature Coax Cable, Black Brilliance VideoFLEX	List 0.42 foot	**$0.35 foot**		Enter Qty foot	IN STOCK
Earn TechCash	**BE-1855AE** Miniature Coax Cable, Black Brilliance VideoFLEX		18 lbs/spool/1000ft **$299.00 spool/1000ft**		Enter Qty spool/1000ft	IN STOCK
Earn TechCash	**BE-1855AR** Miniature Coax Cable, Red Brilliance VideoFLEX	List 0.42 foot	**$0.35 foot**		Enter Qty foot	IN STOCK
Earn TechCash	**BE-1855AR** Miniature Coax Cable, Red Brilliance VideoFLEX		18 lbs/spool/1000ft **$299.00 spool/1000ft**		Enter Qty spool/1000ft	IN STOCK
Earn TechCash	**BE-1855AG** Miniature Coax Cable, Green Brilliance VideoFLEX	List 0.42 foot	**$0.35 foot**		Enter Qty foot	IN STOCK
Earn TechCash	**BE-1855AG** Miniature Coax Cable, Green Brilliance VideoFLEX		18 lbs/spool/1000ft **$299.00 spool/1000ft**		Enter Qty spool/1000ft	IN STOCK
Earn TechCash	**BE-1855AL** Miniature Coax Cable, Blue Brilliance VideoFLEX	List 0.42 foot	**$0.35 foot**		Enter Qty foot	IN STOCK
Earn TechCash	**BE-1855AL** Miniature Coax Cable, Blue Brilliance VideoFLEX		18 lbs/spool/1000ft **$299.00 spool/1000ft**		Enter Qty spool/1000ft	IN STOCK
Earn TechCash	**BE-1855AY** Miniature Coax Cable, Yellow Brilliance VideoFLEX	List 0.42 foot	**$0.35 foot**		Enter Qty foot	IN STOCK
Earn TechCash	**BE-1855AY** Miniature Coax Cable, Yellow Brilliance VideoFLEX		18 lbs/spool/1000ft **$299.00 spool/1000ft**		Enter Qty spool/1000ft	IN STOCK

Earn TechCash	**BE-1855AW** Miniature Coax Cable, White Brilliance VideoFLEX	List 0.42 foot	**$0.35 foot**		Enter Qty foot	IN STOCK
Earn TechCash	**BE-1855AW** Miniature Coax Cable, White Brilliance VideoFLEX		18 lbs/spool/1000ft **$299.00 spool/1000ft**		Enter Qty spool/1000ft	IN STOCK

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7787A
7789A

Brilliance Miniature Dual-Shield Coaxial Cable Bundles

7787A

7789A

- Bundle of 3 or 5 "1855A" type coaxial cables (see above).
- Outside diameter only .44" (7787A) or .54" (7789A)
- 23 AWG solid bare copper center conductors
- Dual shield
- Suitable for indoor and outdoor aerial applications.
- UL/NEC CMR, CEC C(UL) CMG. Flame Resistance: UL 1666 Riser, CSA FT4.
- Full Specifications (GIF)
- Connectors available: "F", RCA, & BNC
- Use our 200004 stripper and 300007 crimper.

An excellent coaxial cable bundle for running baseband video, s-video, component video, or UXGA video. Also works great for line-level audio and subwoofer audio. The small size of this cable bundle makes it easy to route and hide. The low loss makes long runs >300ft possible. **Not suitable for raw satellite, UHF/VHF, or CATV signals.**

Earn TechCash	**BE-7787A** Miniature Coax,Bundle/3,Black Brilliance VideoFLEX	List 2.06 foot	**$1.70 foot**		Enter Qty foot	IN STOCK
Earn TechCash	**BE-7787A** Miniature Coax,Bundle/3,Black Brilliance VideoFLEX		47 lbs/spool/500ft **$770.00 spool/500ft**		Enter Qty spool/500ft	IN STOCK
Earn TechCash	**BE-7789A** Miniature Coax,Bundle/5,Black Brilliance VideoFLEX	List 2.99 foot	**$2.47 foot**		Enter Qty foot	IN STOCK
Earn TechCash	**BE-7789A** Miniature Coax,Bundle/5,Black Brilliance VideoFLEX		73 lbs/spool/500ft **$1055.00 spool/500ft**		Enter Qty spool/500ft	IN STOCK

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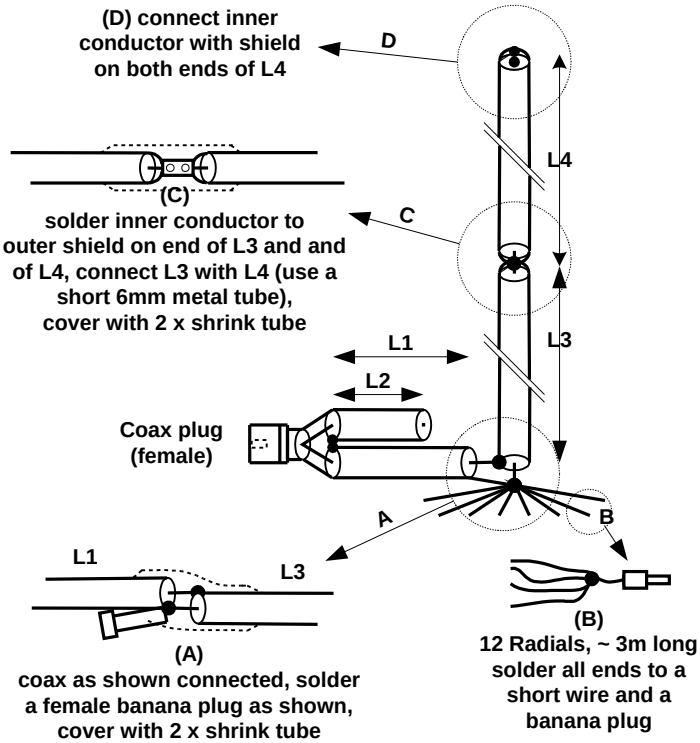
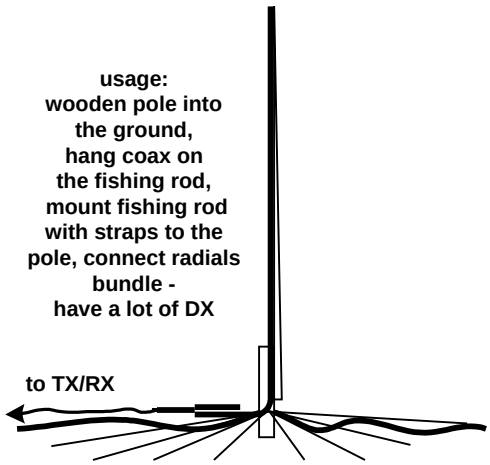
Coax Monopole Antenna
by OE1MWW
Feb. 2002





Feedpoint impedance of antenna is ~ 36 Ohms
YES, it's right, the inner conductor of L1 (feeding coax) is connected to the shield of L3 ! The radiating part of this antenna is the outside shield of L3 and L4.

L1,L2,L3,L3 is made of Coax RG/58 or RG/213

L1 & L2: use the software "coaxstub.exe" to calculate.
`http://www.qsl.net/oe1mww/coaxstub.exe`
you need VB6 runtime!

This document can be found at http://www.qsl.net/oe1mww/coax_monopole.pdf

L3 = f/4 * Vk
L4= (f/4 * 0,95) - L3,
tune L4 for best SWR,
important: inner conductor is connected to shield of L4 on both ends! instead of coax, you can use also a thick wire for L4.

f in meters of wavelength
Vk of coax is ~ 0,66 (RG58U)

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DIRECTIONAL ARRAYS

You have already learned about radiation patterns and directivity of radiation. These topics are important to you because using an antenna with an improper radiation pattern or with the wrong directivity will decrease the overall performance of the system. In the following paragraphs, we discuss in more detail the various types of directional antenna arrays mentioned briefly in the "definition of terms" paragraph above.

Collinear Array

The pattern radiated by the collinear array is similar to that produced by a single dipole. The addition of the second radiator, however, tends to intensify the pattern. Compare the radiation pattern of the dipole (view A of figure 4-25) and the two-element antenna in view B. You will see that each pattern consists of two major lobes in opposite directions along the same axis, QQ1. There is little or no radiation along the PP1 axis. QQ1 represents the line of maximum propagation. You can see that radiation is stronger with an added element. The pattern in view B is sharper, or more directive, than that in view A. This means that the gain along the line of maximum energy propagation is increased and the beam width is decreased. As more elements are added, the effect is heightened, as shown in view C. Unimportant minor lobes are generated as more elements are added.

Figure 4-25. - Single half-wave antenna versus two half-wave antennas in

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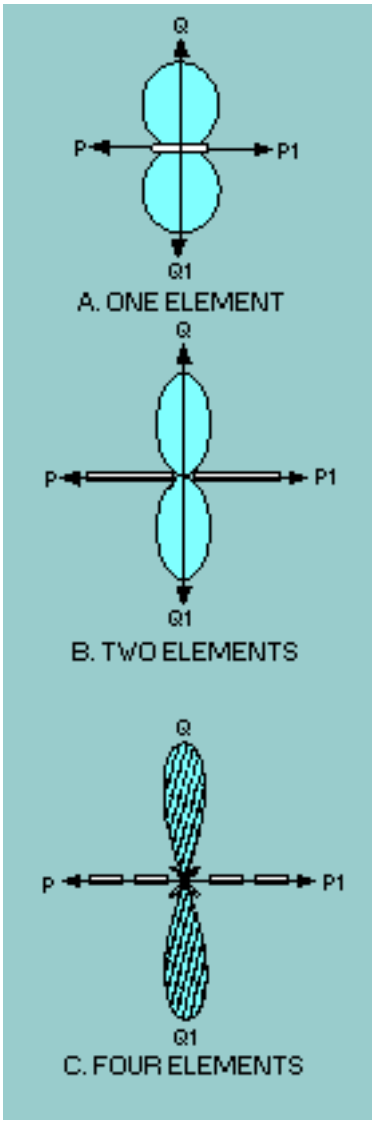
phase.



More than four elements are seldom used because accumulated losses cause the elements farther from the point of feeding to have less current than the nearer ones. This introduces an unbalanced condition in the system and impairs its efficiency. Space limitations often are another reason for restricting the number of elements. Since this type of array is in a single line, rather than in a vertically stacked arrangement, the use of too many elements results in an antenna several wavelengths long.

RADIATION PATTERN. - The characteristic radiation pattern of a given array is obtained at the frequency or band of frequencies at which the system is resonant. The gain and directivity characteristics are lost when the antenna is not used at or near this frequency and the array tunes too sharply. A collinear antenna is more effective than an end-fire array when used off its tuned frequency. This feature is considered when transmission or reception is to be over a wide frequency band. When more

than two elements are used, this advantage largely disappears.

LENGTH AND PHASING. - Although the 1/2 wavelength is the basis for the collinear element, you will find that greater lengths are often used. Effective arrays of this type have been constructed in which the elements are 0.7 and even 0.8 wavelength long. This type of array provides efficient operation at more than one frequency or over a wider frequency range. Whatever length is decided upon, all of the elements in a particular array should closely adhere to that length. If elements of different lengths are combined, current phasing and distribution are changed, throwing the system out of balance and seriously affecting the radiation pattern.

Q.28 What is the maximum number of elements ordinarily used in a collinear array? Answer

Q.29 Why is the number of elements used in a collinear array limited? Answer

Q.30 How can the frequency range of a colinear array be increased? Answer

Q.31 How is directivity of a collinear array affected when the number of elements is increased? Answer

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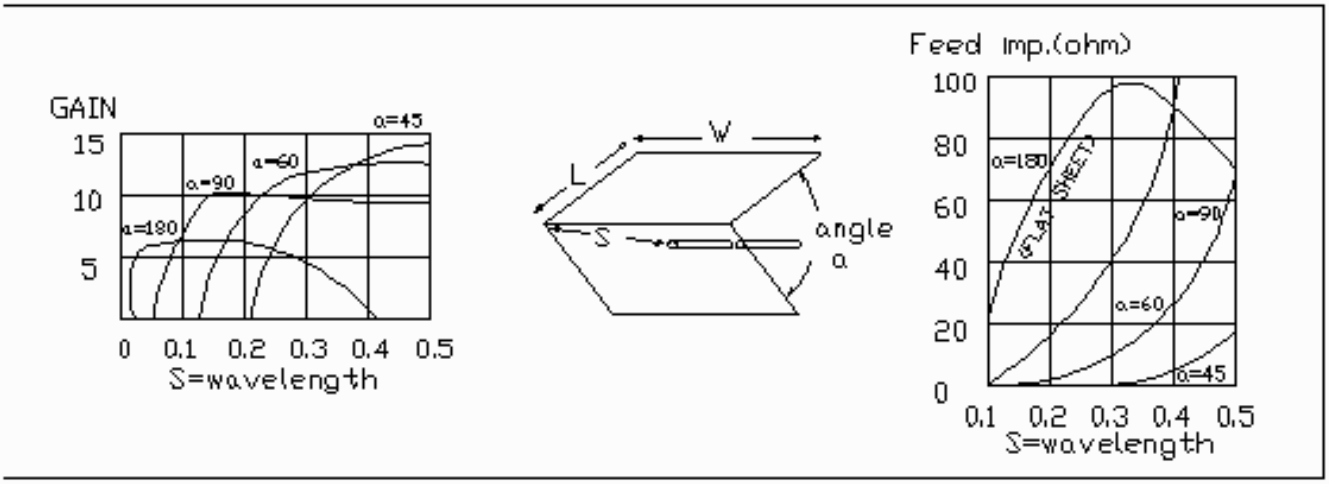
Corner reflector antenna

Do you need a high gain antenna? Have you suffered on picking up interference from unwanted direction? You need a directional antenna but a 12 element Yagi will be too attractive! Well, following might be the answer - a corner antenna. It can provide a forward gain of about 12dbi with a front to back ratio of well over 20dbi.

This design is a periodic plane spaced behind a radiating dipole. The critical factors are the corner angle and the spacing between dipole/vertex (fold point of reflector). The curves in fig.A show that as angle is reduced, the gain becomes progressively greater.

FIG-A RELATIONSHIP BETWEEN ANGLE/SPACING/GAIN



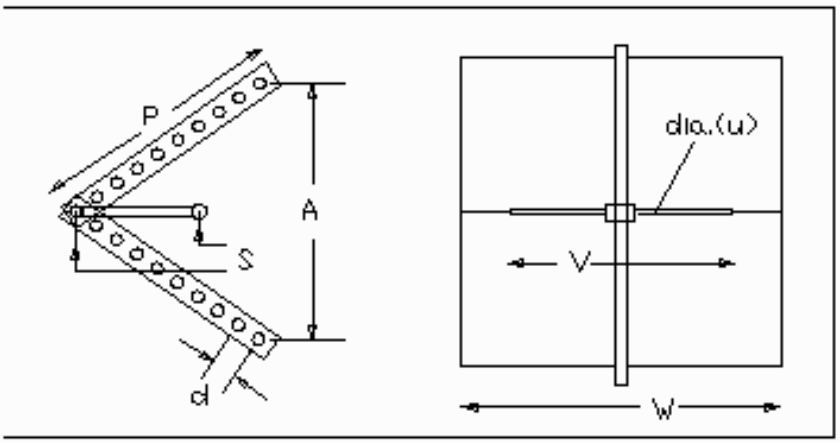


Fig.B
Dimensions for
a 60 degree
reflector antenna

At the same time the feed impedance of the antenna falls towards a lower value and starts creating difficulty in matching. In practice this angle is usually at 90 degree or 60 degree while 90 degree is easier to be matched although gain is lower.

Following are some key points when designing such antenna :

- Length (L) of the sides of reflector should exceed 2x wave-length to secure the characteristics. Reflector width W should be greater then 1x wave-length for a half-wave radiator.
- The reflector can be made of wire netting, sheet metal or even fabricated metal spines arranged in a V-formation. Such spines must be parallel to the radiator with a spine spacing of less then 0.1 wave-length of the operating frequency.
- Spacing between radiator and vertex should be adjustable. This might be the final key to tune-up such an antenna after radiator length is settled for a specific operating frequency.

Impedance of these antennas will change upon operating frequency. Typical value will be around 50 ohm to 75 ohm. A slightly higher S.W.R (1.7:1) has to be expected on lower end of the band. Fig.A also shows relationship between resultant impedance and a change in wavelength (frequency). Following is a table that shows the general dimension of such antenna at UHF and VHF band (Fig.B). All value below are in inches (except **Band** in MHz). Final dimension might vary due to difference in materials employed. Try by error is the key for success.

Band	**p**	**s**	**d**	**v**	**w**	**A**	**u**
144	100	40.0	6	38	50	100	0.375
440	34	13.0	1.4	13	20	34	0.25
435	35	13.5	1.5	13	20	35	0.25

We can discuss it further on the air if you are interested.

By VE3RGW Sept 96.

Return to Dreamland

THE CROSSED-FIELD-ANTENNA
PART I

by
Maurice C. Hately GM3HAT
& Ted Hart W5QJR

PREFACE: Please be aware this article introduces a completely new concept in antenna theory. In fact, this is considered the most important development relative to antennas in this century. It is so pervasive that a series of articles are required to cover the concept and its applications to the depth and extent it deserves. This is the first publication about the ***Crossed-Field Antenna*** (CFA) in the U.S.A. Even as you read this, there are continuing developments in the application of this concept. ***antenneX*** has more than twenty articles about the CFA, including four construction articles (complemented by several more building experiments by readers from around the globe), plus a look at the actual CFA 1992 and 1995 versions of Tanta Station operated in Egypt. This is followed up by a tour of 10 of Dr. Kabbary's 17 more recent CFA broadcast stations in operation in Egypt and Australia. Further, of interest is that the Isle of Man, UK has chosen to install a CFA longwave broadcast station rather than a 845-foot (260m) tower. More stations are going up in Germany, Italy and Brazil. The articles are in the Reading Rooms now. To read more, goto: **Subscribe**

Where will this concept lead ? Let me tell you about one example, lest you think this is all for naught: A CFA only 21 feet (6.5 Meters) tall, located at Tanta in the center of Egypt, provides AM broadcast service at 1.16 MHz (258 Meters) to millions of people from Cairo north to the coast. Certified measurements provide evidence that this small CFA produces a radiated signal almost 6 dB stronger than the previous 1/4 wavelength vertical broadcast tower which was 211 feet (65 Meters) tall. To express the performance a different way >> with the tall tower, a 100,000 watt transmitter was required for the desired coverage. With the miniature CFA the same coverage was attained with the transmitter power reduced to 30,000 watts.

BANDWIDTH GREATER THAN TALL TOWER!

Unlike conventional small antennas, the CFA has greater bandwidth than the tall tower, thus giving better fidelity to the broadcast signal. If that don't tug your heartstrings, you need to go back to making mud pies. For all the rest—you ain't seen nothing yet. In spite of this, the

antenna "experts" have not widely accepted this concept. I can only conclude that those who did not invent it don't want to admit that someone else did. That is a common malady and is often referred to as the not-invented-here (NIH) syndrome. I personally went through the same thing with the small loop antenna. It is the purpose of ***antenneX*** to expose you to new concepts—whether you accept them or not is your personal preference. Some people still don't think man has set foot on the moon!

INTRODUCTION

While thinking about how to present this article, I (W5QJR) was scanning the web and found the following definition of an antenna: ***"An antenna can be any conductive structure that can carry an electrical current.*** **If it carries a time varying electrical current, it will radiate an electromagnetic wave."** In particular note that this definition of an antenna includes a ***current carrying conductor*****.** This is a concept that came about after Marconi threw up a piece of wire and began radio communications. Ever since it has been the accepted traditional concept of an antenna. While it does satisfy fundamental theorem, it is not the only way to develop electromagnetic waves, and therefore is not the only way to build an antenna. In fact, this is a very inefficient way to develop electromagnetic waves.

A number of years ago (1984) I (Ted Hart) wrote the book on small loop antennas and declared in that book, and several related articles, that the only way to build a small high efficiency antenna was to build a small loop. That statement was true until about 1988.

I am simply telling you that one doesn't have to have a wire antenna—there is another way. The other way came about due to a College Professor by the name of ***Maurice Hately, GM3HAT***, in Scotland, and a student named ***Fathi Kabbary*** who came from Egypt to study under Professor Hately while working on his Ph.D. Together, they reviewed the fundamental theory of antennas and decided there was a better way. This occurred about 1988. Subsequently they published articles and patented their new antenna design. Patent 2,215,524 was issued in Great Britain, 626,210 in Australia and others issued in Europe and Japan. In 1992 the U.S. issued Patent Number 5,155,495. But, I am getting ahead of history, so let me take you back to an earlier time.

However, before giving you a history lesson, I do want to note that Mr. Hately and Mr. Kabbary jointly own all rights to this invention. All rights are reserved by them. If you have a commercial interest, contact the author. Mr. Hately reviewed, modified and approved this article, so you are getting this information from the horse's mouth. I, W5QJR, am merely a ghostwriter and take no credit for any of the development. I have derived great pleasure in working with Mr. Hately for several years and he is now in the process of bringing antennas based on this concept for sale thru ***antenneX***.

A HISTORY LESSON

A long time ago, in a far away place, a fellow named Ampere and another fellow named

Faraday (1791-1867) independently developed the concepts and equations to define electromagnetism. Then along came a fellow by the name of Maxwell (1831-1879) who combined the works of Ampere and Faraday and developed the four (4) basic laws called ***Maxwell's Equations***. These equations are so fundamental that every Engineer considers them to be ***the*** four chapters of the engineering bible. Then along came Mr. Heavyside who developed the Differential form of Maxwell's Equations. Somewhere in there, Mr. Poynting presented his equation defining electromagnetic radiation.

Although many authors have quoted Maxwell, and many a college student has struggled trying to understand Maxwell, nothing interesting happened until along came Professor Hately and Mr. Kabbary (now Dr. Kabbary due to this work). As is prudent to do when tying to understand a concept, you go to the most fundamental version of the concept that can be found. In this case, Maxwell's Equations were only an arm's reach away in any good engineering reference book (for example see page 45-4 of ***Reference Data for Radio Engineers***).

Now, our time clock moves forward to March 1989, when ***Electronics and Wireless World*** published an article entitled "Maxwell's Equations and the Crossed-Field Antenna", by F.M. Kabbary, M.C. Hately and B.G. Stewart. To ensure you get all of this picture in proper perspective, the following paraphrases that article. It may get a little technical for some, but that is the essence of this new antenna concept. We will endeavor to translate to simple terms wherever possible, so please continue to read even if you are not an engineer. The picture would not be complete if the theoretical portion of this puzzle was not included in this series of articles.

A LOT OF HEAVY THEORY

All electrical and communications engineers are in some way acquainted with Heavyside's differential form of the third and fourth Maxwell equations, viz.

$$1)\ \Delta \times E = -B'$$
$$2)\ \Delta \times H = J + D'$$

In these equations, **'** is the derivative with respect to time, **E** represents the electric field strength, **H** magnetic field strength, **J** current density, **B** magnetic flux density = μ **H**, and **D** electric displacement = εE. D' is called the displacement current. Equation 1) is Faraday's law, while equation 2) is credited to Maxwell for adding **D'** to Ampere's law, which is $\Delta \times H = J$ to maintain charge conservation or charge continuity and thus obtain **J + D'** as the true or total current.

Unfortunately, the understanding of these equations still poses many conceptual difficulties for many people which inevitably lead to shortcomings in the basic understanding of their

engineering applications (and you thought you were alone?). One reason for this lack of insight is perhaps the inability to appreciate the physical meaning of the vector operations curl, div and grad. Many texts and research papers often detail the mathematical intricacies of these vector operations but few describe in practical simple terms their physical interpretation.

In addition to the above, it is often not realized that contained in equations 1) and 2) is the following extremely valuable information: (a) a time-varying magnetic field creates an electric field (or back EMF) and, importantly, (b) a current ***or*** a time-varying electric field ***or*** both will create a magnetic field. **Please read again and note that you do not need to run current thru a wire to develop a magnetic field.**

The essence of Maxwell's equations, conveyed through points (a) and (b), is that fundamentally they are reactive or field-production equations. The physical , mathematical and engineering importance of the field-production nature may be more readily relayed and understood if the forms of equations 1) and 2) are reversed:

$$3)\ B' \Rightarrow -\Delta \times E$$
$$4)\ J + D' \Rightarrow \Delta \times H$$

The reversal leads not only to a greater understanding of Maxwell's equations (which is hidden in the non-reversed form) but to a greater appreciation of the nature of time-varying electromagnetics and their associated engineering applications.

One significant engineering application, only fully realized through the reversed form of Maxwell's 4th equation, has been the recent development of revolutionary antenna systems called crossed-field-antennas (CFA) which synthesize directly the Poynting vector **S = E X H** from separately stimulated **E** (electric) and **H** (magnetic) fields. **S** is electromagnetic radiation, thus this says there are two (2) components to the radiated field, **E** and **H**. The **X** is defined as the cross product, meaning that they must be properly related both in time, phase, and position. *In other words, if you can separately create the two fields and properly combine them, you don't have to have a piece of wire carrying a current*. Because of this, **A fundamental feature of these antennas is that the physical size of the structure is small and also independent of the radiated wavelength, a truly remarkable concept in relation to present day antenna theory and design techniques.** (Frame that and hang it on the wall).

The principle of Faraday's Law (equation 1) as detailed by most textbooks, is that an electric field can be related to the rate of change of a magnetic field. This electromagnetic feature can be expressed in a more elegant and informative way by reversing equation 1) to give $B' \Rightarrow -\Delta \times E$ which is interpreted as a time varying magnetic flux, **B'** creating an electric field **E** such that the negative of the curl of the induced **E** field distribution is equal to the source **B'**. The directive arrow is present in the relationship to indicate that the left-hand-side causes or

creates the right-hand-side. The negative sign is a manifestation of Lenz's law. In fact, the application of the reversed form of Faraday's law is fully deployed in transformer theory, where a time varying magnetic flux creates, i.e., induces, a back EMF. Note that the **E** field in the reversed form of Faraday's law is the induced **E** field from **B'** and is not in anyway related to the independent electric field created from free charge through Gausse's law.

Consider now equation 2). In magnetostatics, it has always been accepted that current produces a magnetic field though the phenomenon called Ampere's law. To get across the importance of this statement in a more meaningful physical and mathematical form, Ampere's law should be expressed as: $5)\ J \Rightarrow \Delta \times H$ i.e., **J** (current density) creates a magnetic field **H**, such that the curl of **H** is equal to the source **J.** It is also known (though often ignored) that a magnetic field may be related to either a current as above ***or*** a time varying electric field. The latter source of magnetic field is sometimes refereed to as the Maxwell Law, and may be expressed in the more informative form as: $6)\ D' \Rightarrow \Delta \times H$ i.e., displacement current **D'** (a time varying **D** field) creates a magnetic field **H** such that the curl of the H field distribution is equal to the source **D'**. We now see the importance of reversing equation 2) to give equation 4), i.e., **J + D' => D X H,** which should now be interpreted as **J + D'** or both can create a magnetic field **H** such that the curl of the **H** field distribution is equal to the source **J + D'**. The plus sign can, and should, be interpreted as analogous to the digital-logic **OR** symbol.

Unfortunately, many people fail to realize that an **H** field may, at any time, be the combination of two separately induced fields from independent types of sources, i.e., charge motion and displacement current.

The editor says we are out of space, so we will pick up here next month and apply what we have learned to create a magnetic field without running current thru a wire. This will be accomplished by a simple demonstration. That is essential to the process of building this type of Crossed-Field-Antenna. The heavy theory is behind us so it is down hill from here. Stay tuned. **-30-**

Editor's Note: For your convenience, we have placed one of our new Interactive NotePads below for any questions or thoughts while fresh on your mind. Just type in your message (including E-mail address if a reply is needed) in the NotePad and press the Submit button.

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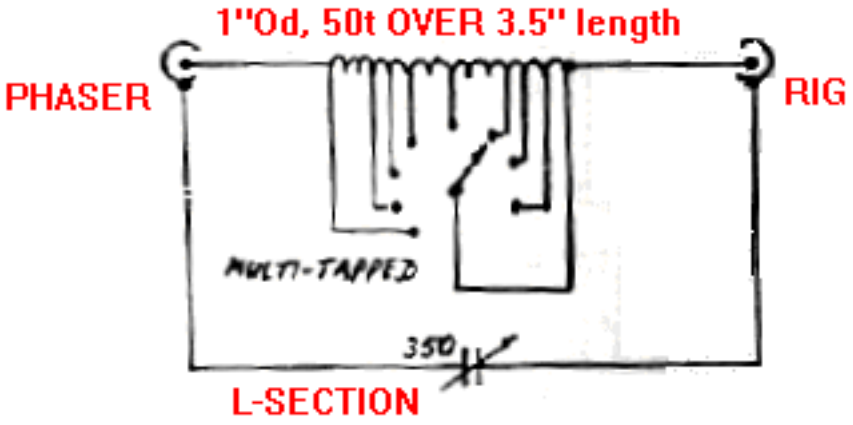
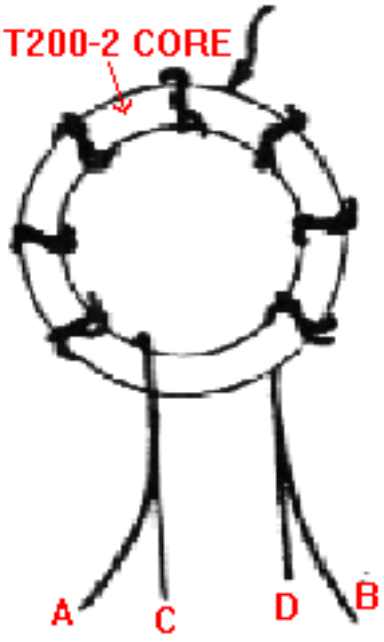
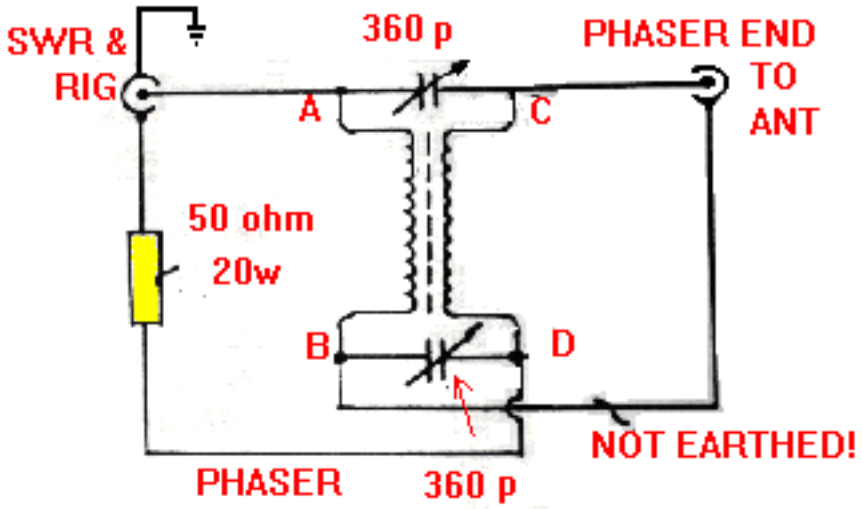
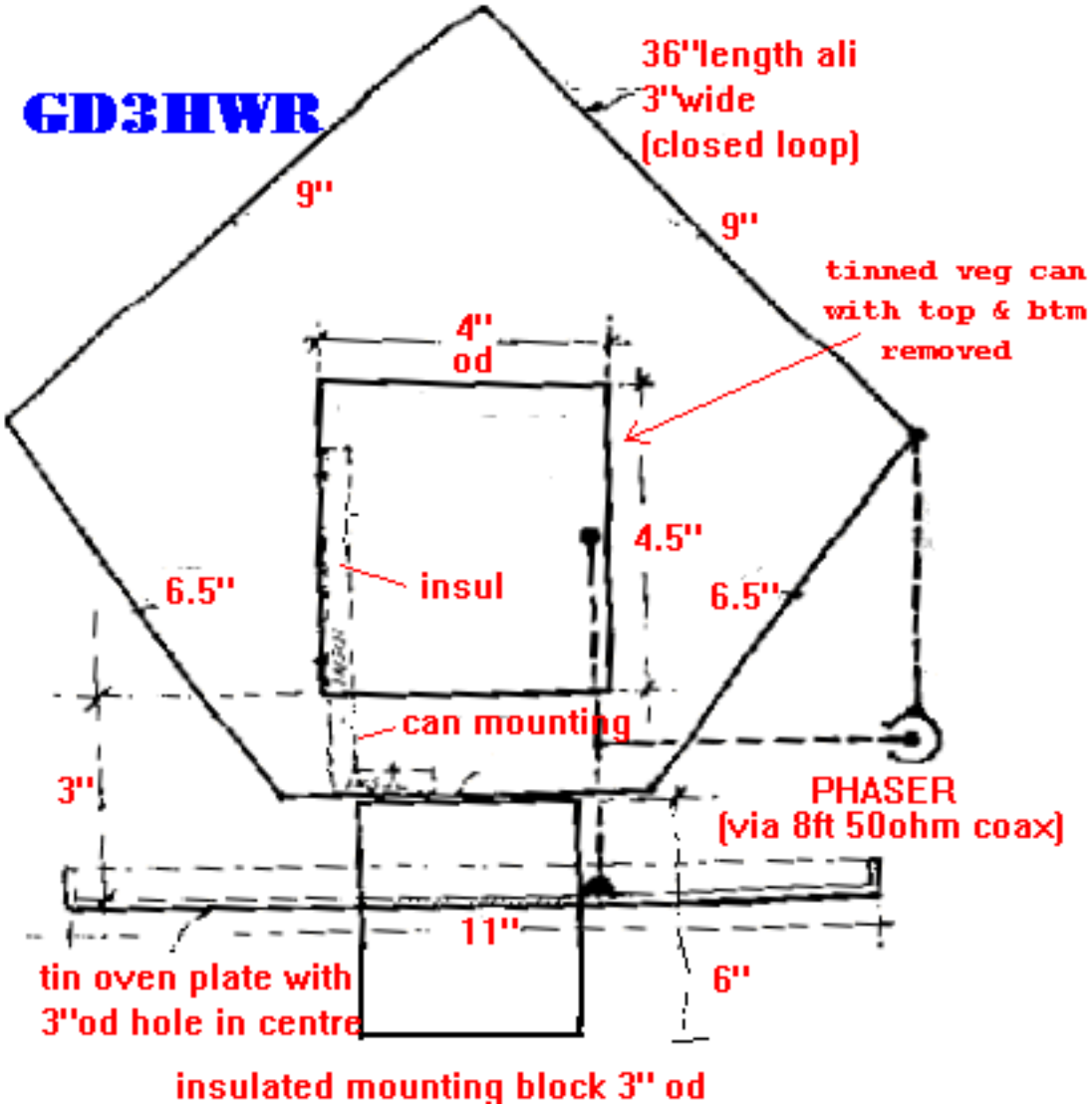
THE MARK II CROSSFIELD ANTENNA









The mark 2 Crossfield Antenna was described in Sprat 76 by **Alec, GD3HQR** . He notes that although it showed no directivity at HF, it was possible to null out time base noise from a TV in an adjacent room. If a valve PA rig with the usual pi section o/p is used, the L-section may not be required. However, it is essential when used with transistor rigs having fixed tuned o/p filters. When tuning the loop, first

approximately tune the L-match by adjusting for maximum RX noise. Then, with QRP, adjust C2 in small steps, at each step adjusting C1 and the L-netwprk for minimum SWR. When this is found, note the settings for future use. Repeat for other bands. Note C2 requires less capacitance at lower frequencies and more at higher frequencies. The author found that, with when properly adjusted, an SWR of at worst 1.5:1 was attained.

It was found that a 1m square sheet of metal - not earthed - placed under the wooden base improved high angle radiation for inter-G working on 3.5/7.0 Mhz.

Note that DC earth is connected to the coax used to connect the L-network to the phaser (it has a capacitive RF earth, which is all that is required).

The author found that his model, used indoors and only six feet above the floor, gave good results on 7 Mhz and above and was 'quite useful' on 3.5 Mhz. It even produced 'the odd QSO' on 1.8 Mhz.

Click to see a photograph of G3FCK's homebrew CFA (108 kilobytes)

(Please note: I have no personal experience of this system and cannot therefore offer any advice or further information. - Frank, G3YCC)

Back to the first page

CUBICAL QUAD ANTENNA CALCULATOR

JavaScript PROGRAM by n6ach@comcast.net

thanks to Scott McClements for the round off problem and to Roger KC5LCA for stream-lining the interface

Bye Mir...thanks for the contacts

Translate this page automatically.

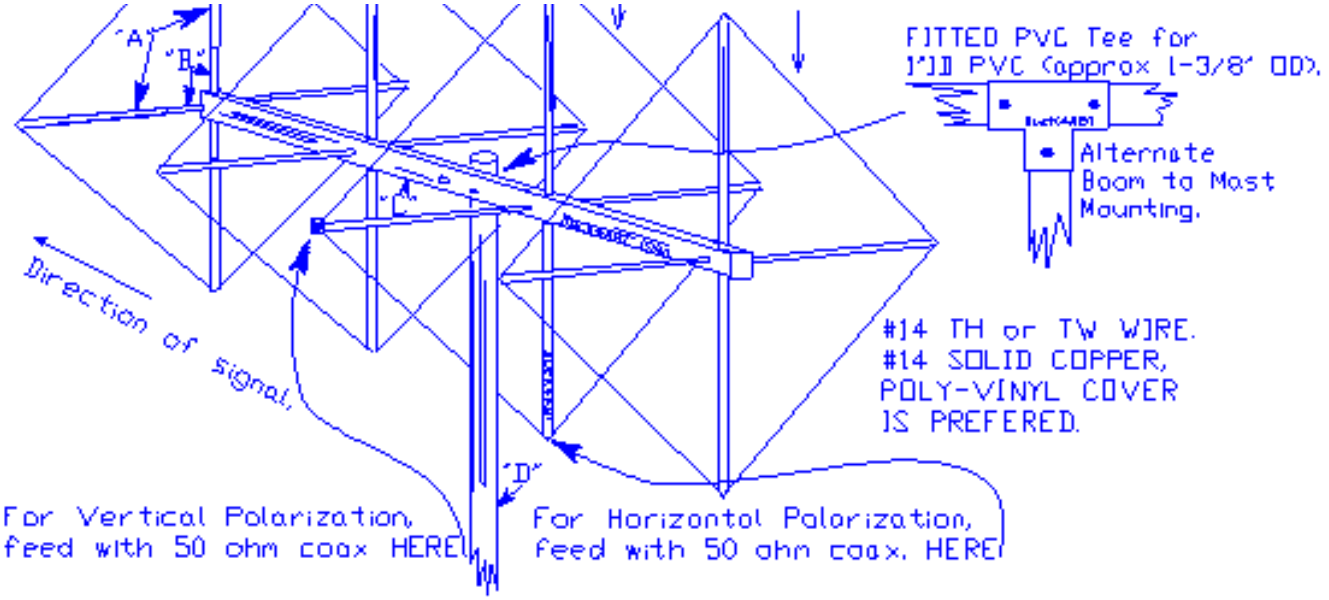
Españoles Les français Deutscher Italiano Portuguêses Japanese Korean Chinese

---

Enter the center frequency: mHz

Elements	Element - Total Length	Wire - One Side	Spreader - Half Length	Spreader - Total Length	Element - Spacing
	INCHES / CM	INCHES / CM	INCHES / CM	INCHES / CM	INCHES / CM
Reflector					
Driven					
Director 1					
Director 2					
Director 3					
Director 4					
Director 5					

Total Boom Length > FEET= INCHES= CM=



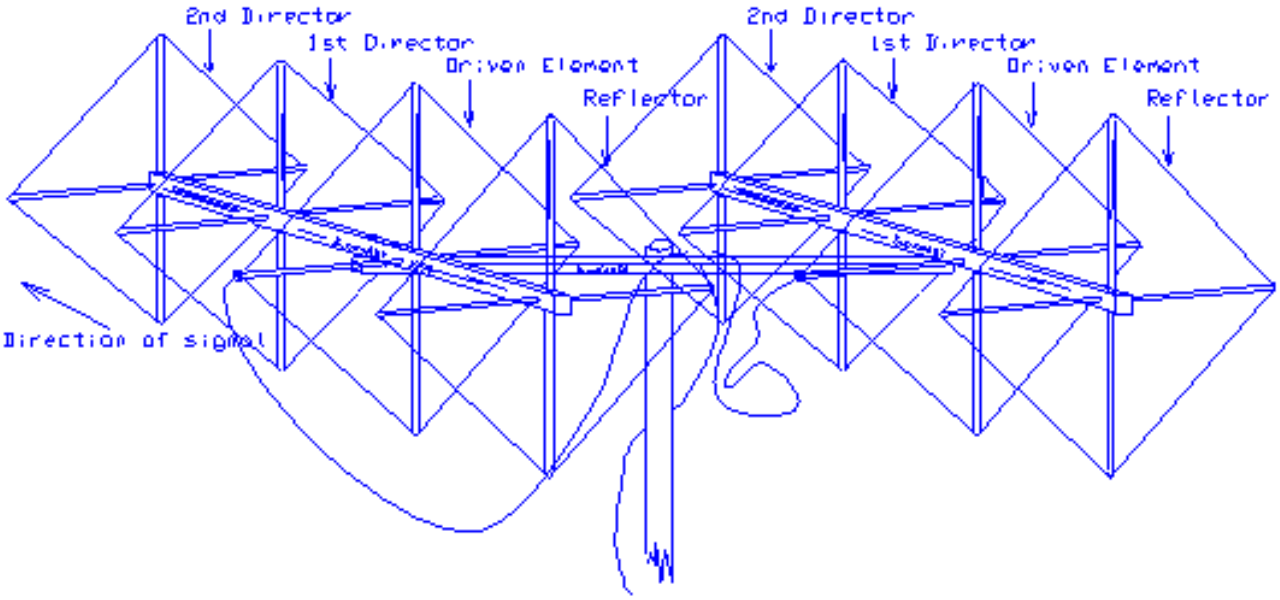
"A" = 1/2" ID hot/cold PVC. PVC should fit securely over 1/2" OD wooden dowel "B."

"B" = 1/2" OD wooden dowel 2 ft long through 1" ID boom. Dowel should extend approximately 9-1/2" out from each side of 1" ID boom. Fit 1/2" PVC ("A") over dowel. Make sure spreader is long enough to support each wire element length as determined by the calculations below. Once PVC is in place over the wooden dowels, drill with small bit and use short wire or picture hanger nail to secure on the dowel. Hot glue may be used as an additional keeper over the wire or nail head.

"C" = Boom may be made from square fiberglass or from round PVC 1" ID. Boom is drilled with 1/2" bit at points where wooden dowels are to be inserted.

NOTE: Make sure the vertical and horizontal dowel holes are drilled offset so the dowels will by-pass each other inside the boom. Drill dowels with 1/16" (or smaller) bit each side of boom and secure with small nail or wire to hold (centered) in place. BOOM is thick wall "Sked 40" 1" ID PVC.

"D" = 1" ID (sked 40/Thick wall) PVC 48" long. Use "fitted" wooden dowel inside to make it more rigid and provide "body" to area where "U" bolts may be attached (rotor head and to boom). See alternate boom mounting.



Horizontal "stacking" of quads will improve the signal strength by at least 3 DB.

Drawing used with permission by *© 0110 1996-98 G. E. "Buck" Rogers Sr.*

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If you have any hints, or comments, please e-mail me at n6ach@comcast.net .
73's de Ron (1Peter 3:15)

03/19/01

(which means back)

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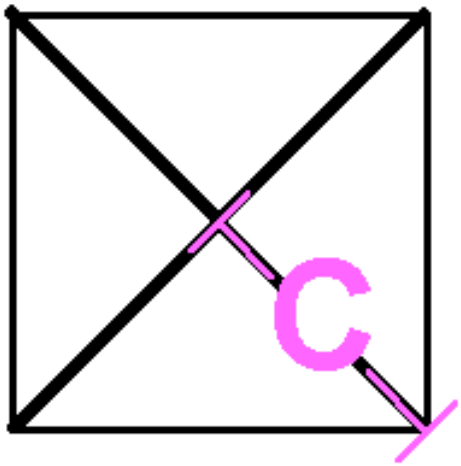
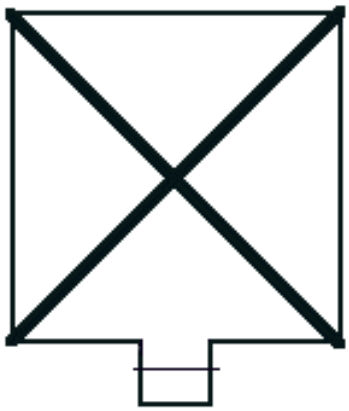


C= Spreader arm Length



Reflector tuning stub

How to mount and feed.

Horizontal or vertical polarization of an cubical QUAD-Antenna is determined by placement of the feed point (feed gap on the driven element) - Parasitic elements need to be mounted in the same orientation as the driven element. If tuning gap is present on the parasitic elements, doesn't the placement of this gap effect the polarization. Feed line is made out of standard coaxial cable (RG8/RG58) 50ohm unbalanced, to attach this feed line to the balanced drive element (Quad loop) without proper matching is not advisable, instead use an 1:1 balun
see W2FMI High Power Baluns

Gain vs Spacing

The gain of an antenna with parasitic elements varies with the spacing and tuning of the elements, and thus for any given spacing there is a tuning condition that will give maximum gain this spacing. The maximum front to back ratio seldom, if ever occurs at the same condition that gives maximum forward gain. The impedance of the driven element also varies with the tuning and spacing, and thus the antenna system must be tuned to its final condition before the match between the coax and the antenna can be completed. However, the tuning and matching may interlock to some extent, and it is usually necessary to run through the adjustments several times to insure that the best possible tuning has been obtained.

Order *How to Build and Adjust Quads*

or

The Quad Antenna: A Comprehensive Guide to the Construction, Design, and Performance of Quad Antennas

For some great construction information stop in to ***Scott's quad page***

Cubical Quad Antenna Design:

By: Dr. Carl O. Jelinek N6VNG

The frequency scaling formulas for Cubical Quad antennas are shown in Figure 1. Note that frequency is measured in Megahertz {MHz} and the total length of each element is measured in feet {ft}. The spacing of each element is the same and all directors are the same size. The gamma match uses a small air variable capacitor approximately the value given and an adjustable shorting bar at the end connected to the element. The antenna is tuned by adjusting the length of the shorting bar on the gamma match for minimum VSWR with the variable capacitor half engaged. Then adjust the capacitor for minimum VSWR at the mid band frequency. The enclosed EXCEL program can be used to determine the element lengths and the gamma match values for different frequencies. In general the design is robust and may be optimized for gain or front to back by adjusting the spacing of the elements. This antenna design has been built for both the ten and two meter versions and I have used them for T-Hunting and in two CVARC Field Day events with good results. Figures 2 and 3 show the calculated antenna patterns and performance. Table 1 shows an example output from the EXCEL Scaling program. Figure 4. Shows VSWR vs. Frequency for three different 2 meter antennas, a 4 element Quad, a 6 and a 12 element Yagi.

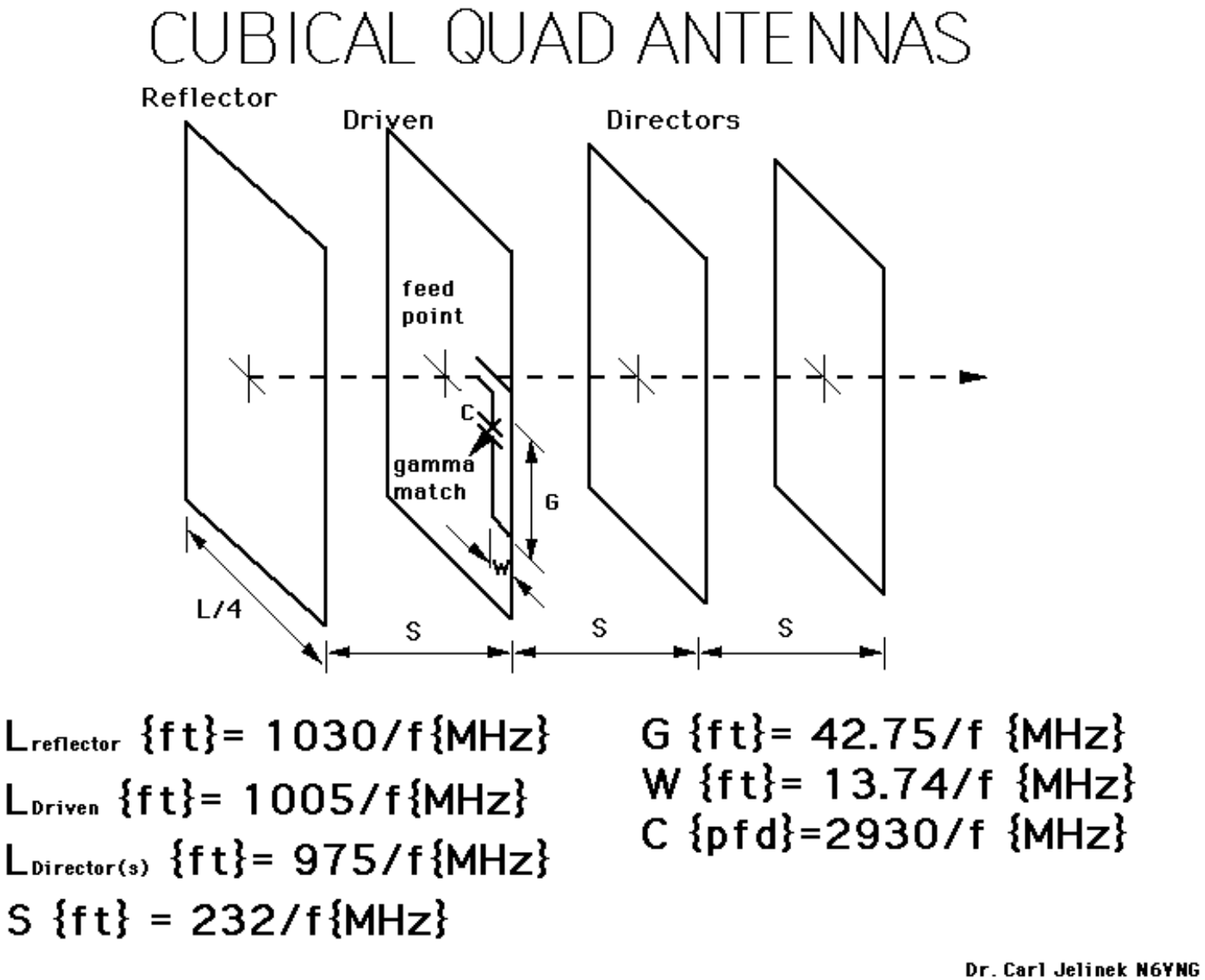


Figure 1. Cubical Quad Frequency Scaling Equations.

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Figure 2. Elevation Pattern for the 4 element Cubical Quad 36" above a perfect ground

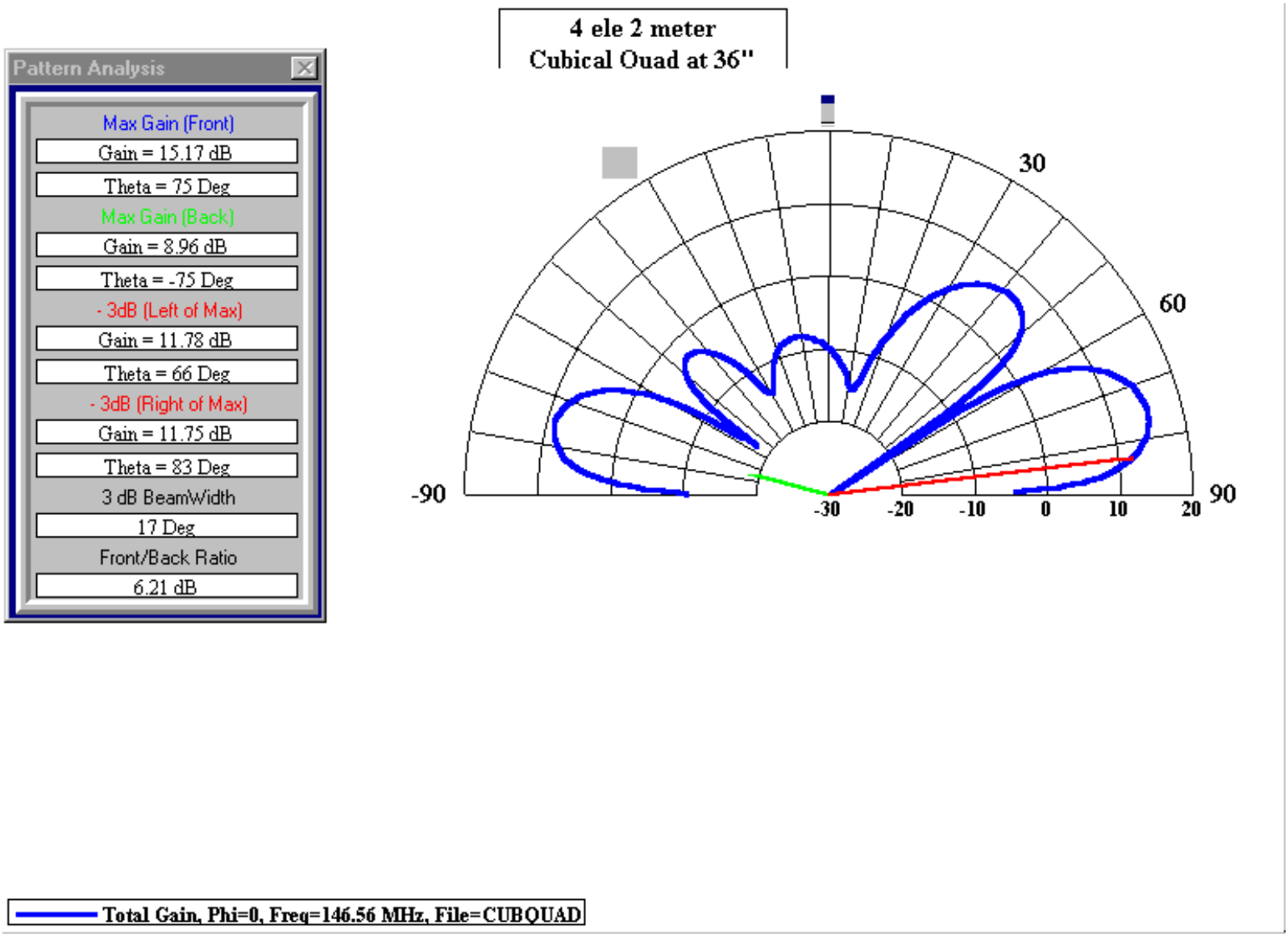

Figure 3. Azimuth Pattern for the 4 element Cubical Quad 36" above a perfect ground

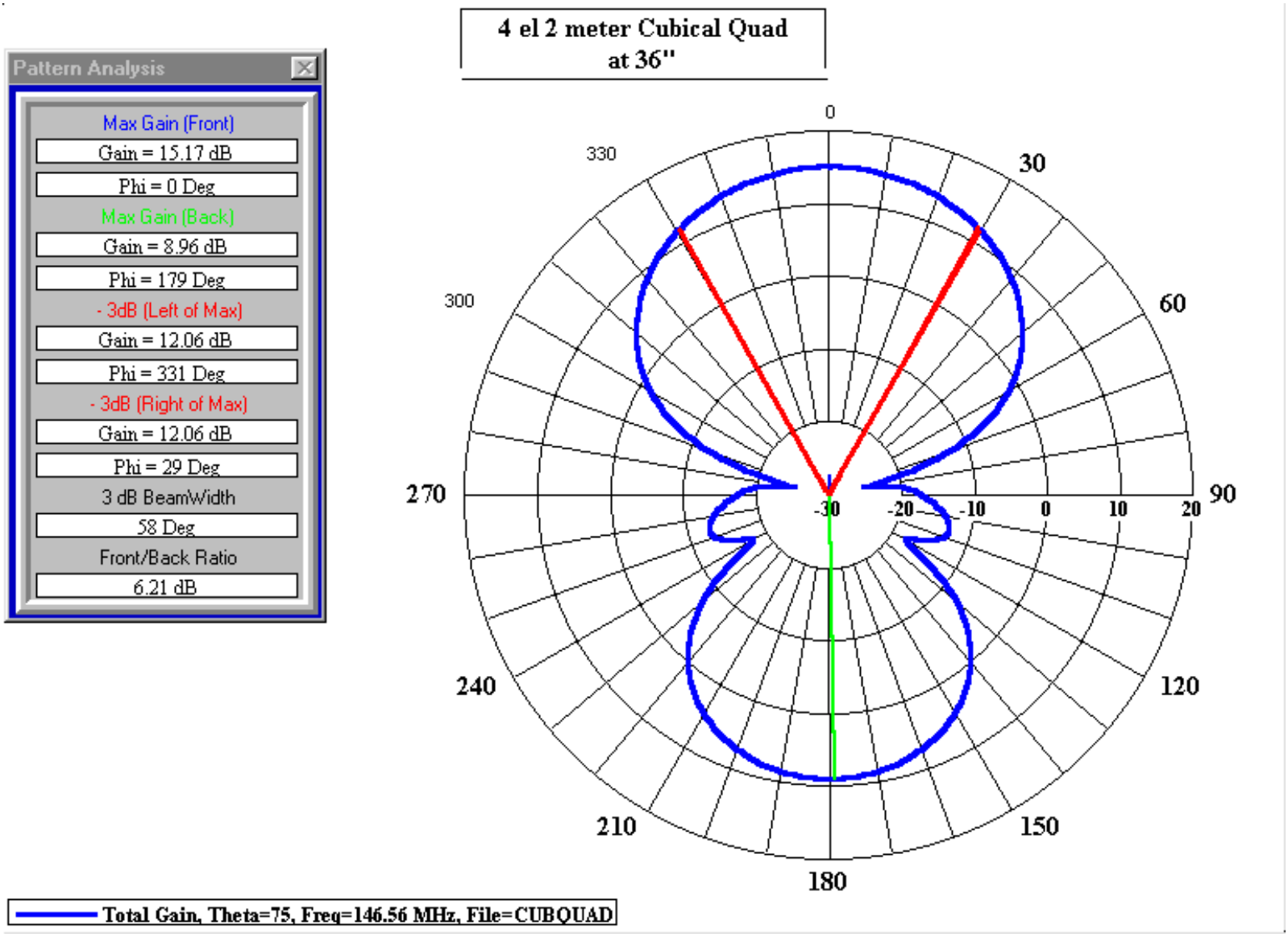

Table 1. Cubical Quad Scaling Relationships

CUBICAL QUAD	Scaling Formulas		3/19/97 22:24				
Dr. Carl O. Jelinek	Total Length		All the same	All the same			
N6VNG	Lr	Ldrv	Ldir	S	G	W	C
f=Frequency {MHz}	Reflector {ft}	Driven {ft}	Directors {ft}	Spacing {ft}	Gama {ft}	Width {ft}	Cap {pfd}
146.565	7.028	6.857	6.652	1.584	0.292	0.094	20.0
222	4.640	4.527	4.392	1.045	0.193	0.062	13.2
445	2.315	2.258	2.191	0.522	0.096	0.031	6.6
52	19.808	19.327	18.750	4.463	0.822	0.264	56.3
28.5	36.140	35.263	34.211	8.144	1.500	0.482	102.8
Notes: Scaling	Lr = 1030/f	Ldrv = 1005/f	Ldir = 975/f	S = 232.1/f	G = 42.75/f	W = 13.74/f	C = 2930/f

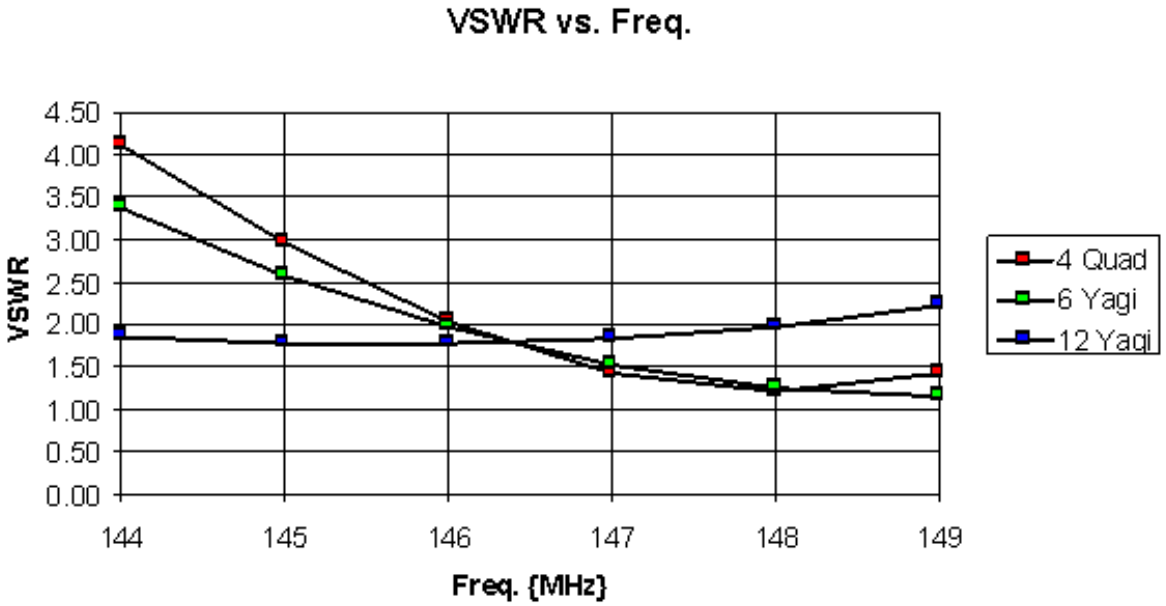

Figure 4. VSWR vs. Frequency for Three 2 Meter Antenna Designs.

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Cubic Quads

KQ6RH

(C) 1998, 1999, 2000

Ray Jurgens

(Up-Dated 2/25/2000)

Cubic Quads

The cubic quad is a very popular way to get reasonably high gain and excellent front to back ratios as well as low angles of radiation for without going to extreme heights. Here I present several designs that that achieve the great performance that hams have associated with this antenna for years. Data are presented for 2 and 6 meter quads and a combined 2 and 6 meter quad that is optimized. The 2 meter 3 element design gets a great 9.5 dBi gain coupled with a F/B ratio of 23 dB.

Light weight portable cubic quads can be constructed rather easily from fiberglass tubes supported by central hubs. You should be familiar with the material presented in the Quad Loop and Pfeiffer Quad sections of the Antenna Magic page. Cubic quads for wavelengths shorter than 15 meters are easily constructed, however, you should be aware that the weight of these structures is larger by a factor of about 3 relative to most of the planar designs presented in the main menu. Because of this, a heavier mast must be used to support the structure in most cases. Also, be aware that the space needed to assemble and raise a full cubic quad is larger than for the planar designs, and this may be a significant limitation imposed in some locations. In my own case, the backyard associated with my town house is barely large enough to assemble a cubic quad with spreaders of 8' in length. Wires and guy cords get tangled in the fruit trees, and spreaders hang over into neighbor's yards. Anything larger 8' with extended spreaders is essentially impossible to assemble without working above the level of the fence and fruit trees. For that reason, I shall present only two designs which are more or less typical of what can be done easily. The two designs presented are for HF and VHF and should be useful to a wide audience. The HF design is a two element quad for 10, 12, and 15 meters while the VHF design is a two element design for 6 meters with three elements for two meters. A specific advantage of the standard quad design is that multi-band operation is easily accommodated.

2 and 6 Meter Quad

The spreader length necessary to support a 6 meter quad is less than 3.5 feet, so the standard 8' lengths of fiberglass tubing can be cut in half to make 4' sections. It is also possible to telescope shorter sections of 1/2" and 1/4" tubing to make a slightly lighter weight design. In that case, the 8' sections could be cut in quarters and the overlap of about a half a foot would be entirely adequate for the telescoping leaving spreaders of about 3.5 feet. The boom length for a full quarter wave spacing is less than 4.75 feet, so a single 8' piece of 1" fiberglass tubing is more than ample.

Looking quickly at the 2 meter requirements, the spreaders need be no longer than 1.25 feet and a three element wide spaced boom requires no more than 31.25", thus this can be easily tucked between the 6 meter 2 element quad. In fact, it is necessary to stretch it out a bit. So, a common design requires beginning with an optimized design for 6 meters and accommodating the 2 meter design to the locations of its two hubs. The third 2 meter hub occupies a space between the two 6 meter hubs.

Looking at an optimized 2-element 6-meter design, the following parameters give excellent performance:

Parameter	Length in "	Length of Loop in "
Reflector Loop	21.42 Side	85.70
Driven Loop	20.60 Side	82.38
Director Loop	20.18 Side	80.72
Reflector Location	-11.77 Boom	Relative to Driven Element
Director Location	19.44 Boom	Relative to Driven Element
Total Boom Length	31.21 Boom	

Table 1

Dimensions of a 3 Element Cubic Quad for 2 Meters

The 6 meter parameters are given in Table 2, which is a simple two element design:

Parameter	Length in "	Length of Loop in "
Reflector Loop	58.91 Side	235,64
Driven Loop	55.87 Side	223.48
Boom Length	47.41 Boom	

Table 2
Parameters for 6 Meter 2 Element Quad

The next step is force the 2 meter 3 element quad to have a total boom length identical to that of the 6 meter quad. Note that the 2 meter 3 element easily fits within the same space as the 6 meter 2 element, that the 2 meter quad will have a longer boom than is considered optimum.

(To Be Continued)

10, 12, and 15 Meter Quad

The construction of 2 element cubic quads for 10, 12, and 15 meters is not very difficult, but the structure requires some guy strings to keep the light weight elements from bending. The bending actually would not degrade the performance very much, but the nice square structure clearly looks better, and it will probably hold up better under high winds. In order to keep the spreaders from bending due to gravity loading and wire loading, you will need to have a place to guy them from two directions. The antenna wires can serve as structural elements to help reduce the bending in the plane of the loops. However, bending is also a problem perpendicular to the plane of the loops, and guy strings are necessary to stabilize that direction as well. The easiest way to provide a place to connect the guys is to use a boom extension. The extension does not need to be any longer than 3 feet for 8' spreaders. In the case of 15-meters, the boom can be about 8', and the two extension increase the boom to 14' or a turning radius of 7'. The actual turning radius of a quad depends upon whether it is set up as a diamond or square configuration, the square being the smaller of the two.

Return to Main Menu

GO TO HOME PAGE

Theory - Practice - my Quad

MY QUAD (2 EL. FOR 10-12-15 MT)

I built two quad (one to me, the other to Mario, ik7zcq).

I used "+ " ("Diamond") configuration which gives just 0,1-0,2 db more in the gain than "Square" configuration: I built it in this way only for mechanical reasons. I used square-aluminum boom 30*30*1 mm 180 cm length; into it I put a steel pipe just to make it stronger; the mast to boom attachment comes from a CB antenna ($6)(see picture 1).

The spreader to boom attachment (one for each element) is home-made using a square-steel bar (35 cm length); on the tips there are 4 perpendicular

pipe (25 mm diameter) where you'll insert poles (see picture 2).

Poles are 5 mt length (21 mhz); I use four bamboo poles for each element ($1 each); each pole is 240 cm length so I made it longer using in the lower part a short aluminum pipe (25 mm diameter, 100 cm length).

I gave the poles a protective coat of boat-paint and then I covered it with PVC ribbon (thanks to Jan, G0KRL for info).
Loops are on copper-insulated multi-wire (1mm for my quad, 3mm for Mario's one).
Here are formulas for loop length:

L(Driven)=7658/F
L(Reflector)=7849/F
L(Director)=7430/F
L=Length for one side.

I never spent time tuning it (it's "no problem, no tune", and works fine!) but I know working some hours I can optimize Gain or F/B Ratio ecc.

I fed driven elements with 3 (one for each band) 1/4 lambda stub using TV 75 Ohm cable; on the Mario's quad i used RG 11 useful for high power; over the outside of the feed line, near the feed point, there are some ferrite beads taped up and sealed against moisture: that's my balun (see picture 3); on the boom there's also the rele' box for antenna switch (see picture 4); then a common 50 ohm RG 213 goes down on the shack.
I suggest you do not solder the stub directly on the loop, but using a SO239: here it's still working fine but will be better how i suggest.

The weigh of my quad is about 15 Kg.

HOW DOES IT WORK?

Oh guys! It works very fine! My quad is only 8 mt up, but it never performed worse than a 3 element yagi (sometime I hear what yagi doesn't; I don't know why european stations give me better reports than mine (If I give them 5/7 they give me 5/9, 5/9+); with Dx station reports are the same. The same occur with F/B Ratio: it's about 15 db with european stations, more with Dx.
Using the quad I forgot my antenna-matcher: SWR is less than 1:1,6 at the end of the band;

I think home-brew this antenna is easy: everybody can do it! I spent less than 100$ and I have a very fine beam; may be this summer I'll add 18 e 14 loops but this is an other story…

73 de iz7ath, Talino

Quad Loops

KQ6RH

(C) 1998, 1999, 2000

Ray Jurgens

(Up-Dated 2/25/2000)

Quad Loops

Quad loops can be a practical way to get some gain without constructing a full Cubic Quad. The free-space gain of a square loop is about 3.13 dB which is a little improvement over a dipole. If you raise the center of the loop up just one half wave length, a peak gain of 7.75 dBi is reached near 24 degrees elevation. This is not too bad for a simple antenna, and even for 20 meters, the center height will only be about 32 feet. You have the choice of either horizontal or vertical polarization and the decision as to whether to set the spreaders up as a cross or and X. Most people choose horizontal polarization which places the feed at the bottom of the loop. The cross or X selection may depend upon the structure and how close the antenna is to the ground. Normally, the X construction can be more stable in that the weight of the wires is supported by two spreaders while a single spreader must support the entire wire mass as well as part of the weight of the horizontal spreaders in the cross configuration. The cross configuration has definite advantages in height over the X and puts more of the wire at a greater height above the ground. This configuration puts a high current region at the upper apex and gives a better elevation pattern for a given height of center above ground. And if you live in lands of snow and ice, the water runs down the wires and off the corners rather then freezing to the horizontal wires. A definite advantage!

I have constructed quad loops for several bands, but most of my effort went toward a 17 meter loop using 1/2" diameter spreaders extended by 1/4" telescoping sections approximately 2 feet long. There are few secrets to success here, however, the information in the ARRL handbook for driven elements seems to give antennas that are a little small, but that could be due to my environment. The ARRL Antenna book gives the formula 1005/fMhz for computing the length of wire needed for the full loop. It may be better to begin with 1030/fMhz and cut back if the frequency comes out too low. The #14 flex-weave wire is a good choice as it stretches out easily, however, it is definitely over-kill and much lighter wire can be used.

The driving impedance of these loops is a little inconvenient, mine measured about 130 Ohms which agrees well with theory, so a 2:1 balun is recommended or a quarter wave of 75 Ohm

line as a matching section. I have found it a bit of a trick to make a good balun that raises the impedance by a factor of two, but there are excellent directions in Jerry Sevick's, W2FMI, book "Building and Using Baluns and Ununs." It is important to make the characteristic impedance of the transmission line winding 100 Ohms as directed, otherwise the high frequency properties will be compromised. An alternative way to feed these loops is to use 300 Ohm twin line or ladder line. I've used a half wave length of 300 Ohm foam filled line with good results by connecting to the balanced line terminals on the antenna tuner. This presents roughly 130 Ohms to the tuner that is expecting 200 Ohms for a perfect match. This small SWR is easily tuned out. And, if weight is an issue, then this is definitely the way to go!

Let's consider the spreader length. Using the formula for the wire length (1005/fMHz), we see that the spreader must be at least (1005/fMHz)/(4*sqrt(2)). For 15 meters, this comes out to about 8.4 feet. Thus, short extensions to the 8' spreaders are required. The extra 0.4' is really no problem for 1/4" rods or tubes, but 17 and 20 meters require special attention. The spreaders for 17 meters must be about 9.82 feet, or an extra 1' 10". In the case of 20 meters, the extensions need to be 4.5'. This latter case is really pressing the limits of what can be done with the light weight materials. All versions of the quad loops require guy lines for wavelengths longer than 15 meters. This means that you must have a boom section that extends about three or four feet either side of the hub. The guy lines can be Nylon fish line, however, Kevlar line works better in this application,because it does not stretch like the Nylon and the lengths can be accurately calculated. Even with guy lines, the 20 meter loop will require lighter wire, i.e., you will not be able to stretch out #14 flex-weave on this structure. Although the 15 meter structure can be managed without guy lines, the spreaders can be bent out to tension the wire. Using #20 stranded copper wire will cost less than 0.25 dB of gain. So you could use stranded hook-up wire. This wire normally comes with insulation, so, you could strip it off exposing the wire to weather, or leave it on and shorten the length by a few percent to compensate for the dielectric loading. I suggest that you leave the insulation on, and buy black Teflon insulation to make the wire less visible. Radio Shack offers a black #18 hook-up wire that also works well.

The construction of loops for 2, 6, 10, 12, and 15 meters is rather straight forward, so the parts list is short. The construction of loops for 17 and 20 meters requires more extensive guy lines, so the parts list grows to cover extra guy ties and spreader extension. We suggest using a short boom with guys for all bands longer in wavelength than 12 meters. As a short boom is required to put the hub in vertical position, adding extra length to the boom and the extra guy lines is really no inconvenience and provides a much more stable structure. Loops for 2 and 6 meters require no guys at all. You should order only enough 1/2" tubing to make the four spreaders for the VHF loops. The 1/2" Guy Ties are convenient for mounting the wire.

In all cases, you will need a mast-to-boom plate with appropriate U-bolts for mounting the 1" boom to the upper mast section. You also must decide if the hub or the boom plate is to be in the center, i.e., both can not be in the center of the boom. In general, it is simpler to put the hub in the center in that all the guy lines can be cut the same. This choice slightly unbalances the the weight at the top of the mast, but this can be compensated easily by adding weight (fishing

sinkers) to the shorter side of the boom relative to the boom plate. The construction of boom plates is given in detail in the ARRL Radio Amateur's Handbook. The material of choice is usually aluminum, however, I have been using 1/4" PVC plates with success, and it is even easier to cut and drill than the aluminum and seems to have adequate strength. I have been using four U-Bolts on both the mast and the boom, i.e., two on each side of the mast and boom.

You should order the following parts for 10 and 12 meters:

Item	Quantity	Description
1	1	HUB 4-050-100, Central Quad Hub (RFJ)
2	4	8' 1/2" OD fiberglass tubing, spreaders (MGS)
3	1	8' 1" OD fiberglass tubing, boom (MGS)
4	4	GT 4-050 1/2" Guy Ties, for tips of spreaders (RFJ)
5	2	GT 4-100 1" Guy Ties, for tips of boom (RFJ)

Additions for 15 meters:

Item	Quantity	Description
6	1	1/4" OD fiberglass tubing, spreader extensions, quarter it! (MGS)
7	4	1/2" stainless steel hose clamps, compression clamps for extensions
8	4	GT 4-025 1/4" Guy Ties, for tips of extensions or wire attachment (RFJ)

Additions for 17 meters:

Item	Quantity	Description
9	1	8' 1/4" OD fiberglass tubing, spreader extension, halve it! (MGS)

Additions for 20 meters:

Item	Quantity	Description
10	2	8' 1/4" OD fiberglass tubes, spreader extenders (MGS)
11	8	GT 4-050 1/2" Guy Ties for cross truss ties, (RFJ)

Note, the spreader extenders can be full 8' lengths of 1/4" tubing, i.e., you don't need to cut them, but you can reduce the weight slightly if you do. They can be cut to 5' lengths leaving 3'

pieces for use with smaller quads. The 20 meter structure requires cross truss guy lines (for details see our Construction Page). These require more guy tie points near the centers of the 1/2" spreaders. It is easier to align the tensions if the Guy Ties are separate for the interior and exterior guys, but common tie points can be used if the guy lengths are cut accurately. The total number of guy lines is as follows:

Item #	Quantity	Approx. Length ft.	Description
1	8	9.0	interior spreader to boom posts (4' post assumed)
2	8	13.2	exterior spreader extension to boom posts
3	4	11.3	interior perimeter guys (could be wire loop)
4	4	17.7	exterior perimeter guys (could be wire loop)
5	8	9.0	interior cross lateral trusses
6	8	13.2	exterior cross lateral trusses

Table of Guy Line Lengths and Definitions

for a grand total of 40 lines requiring 80 attachment clips and a total length of material of 1084.4 ft. You should plan a full day for the first construction of the 20 meter loop. Once assembled, the antenna can be disassembled in about the time required to roll up the guy material. Re-assembly takes about an hour.

Return to Main Menu

PA0FRI

RadioZendAmateur/HAM, Frits Geerligs, QTH: Etten-Leur

PA0FRI FHV Geerligs, A-class NL Licence/CEPT-class morsecode included

Technical Homepage for HAM's

INFO	PAØFRI	HAMS	UPDATED 06sep2004	BIOGRAFIE	LINKS	CONTACT
QRO	ANT	ATU		MODS	MISC	PS

Een aantal onderwerpen met veel foto's zijn ingekort omdat de limiet van 25 MB bij de server bijna bereikt is.

Informatie

Doel

Deze eenvoudige site geeft informatie over mijn experimenten, lopende projecten, modificaties en eerder gepubliceerde ontwerpen. Teksten, schema's, foto's en artikelen worden aangepast of toegevoegd als er nieuwe ontwikkelingen of inzichten zijn. Artikelen mogen elders gepubliceerd worden als u de bron erbij vermeldt.

Reden

Veel heb ik geleerd van artikelen die door andere zendamateurs geschreven zijn in Electron en andere electronica bladen. Als tegenprestatie probeer ik hetzelfde te doen voor anderen.

Doelgroep

Volgens vele zendamateurs wordt tegenwoordig in tijdschriften van verenigingen een onderwerp veel te theoretisch gepresenteerd. Daarom zal hier zoveel mogelijk de praktische kant toegelicht worden. Beginnende of nog onervaren zendamateurs worden dan mogelijk ertoe aangezet om ook eens te experimenteren of zelf iets te maken. U hoeft zich niet te schamen, begin eenvoudig en bedenk dat velen over zelfbouw praten, maar het verder daarbij laten. De meeste kennis wordt opgedaan door het zelf te doen. Tijdens het bouwen en experimenteren wordt inzicht verkregen, vooral als het niet meteen lukt of er iets kapot gaat. Men gaat fout zoeken, meten, vragen, lezen en proberen. Als tenslotte het zelfbouwproject goed werkt of de koopdoos door u is gerepareerd, verbouwd of afgeregeld, dan is de voldoening niet te beschrijven. Er is geen drempelvrees meer en een verlangen naar een volgende uitdaging is een feit.

BACK BACK

Activiteiten PAØFRI

Mijn voornaamste bezigheden zijn experimenteren, bedenken, veranderen, proberen, "ontwerpen" en af en toe publiceren. Er zijn ongeveer 20 verschillende praktische onderwerpen van mijn hand verschenen in ELECTRON (14), CQ-PA, Radio Communication, CQ Friese Wouden, QRP NIEUWSBRIEF en andere tijdschriften die artikelen hebben overgenomen. Zelfbouw wordt ook veel gedaan, maar als alles naar wens werkt neemt de belangstelling af. Vaak wordt het bouwsel weer gedeeltelijk of volledig gedemonteerd om over onderdelen te beschikken voor een ander idee. Als het de bedoeling is om het apparaat te gebruiken, wordt het ook netjes afgebouwd, maar het afwerken van de frontplaat laat nog wel eens lan…g op zich wachten. Veel schakelingen en andere bedenksels die hier getoond worden zijn eigen ontwerpen. Praktische ervaring is vooral opgedaan met antennes, antenne tuners, baluns, HF versterkers, ontvangeringangen, mixers, AF& HF spraak processors en 2m transistor eindtrappen.

Voornamelijk wordt gewerkt op HF met een gemodificeerde Ten Tec Corsair II, R5 vertical en inverted zelfbouw W3DZZ gevoed met openlijn via een S-Match symmetrische antennetuner.

Er zijn zelfbouw lineaire versterkers FRI-400 (4 × PL519 400W), FRI-750 (TB3/750, 750W), een gemodificeerde Ten Tec Centaur (3 × 572B, 800W) en een opgeknapte Heathkit SB-200. In ontwikkeling is een FRI-1500 (1500W) met een Russische tetrode GU-43B in een passief rooster schakeling. Lineairs worden zelden daadwerkelijk ingezet op HF banden. Het bouwen is leuker en zij worden het meest gebruikt om tuners, dymmy-loads, low-pas filters, coaxkabels en openlijn te testen.

Inmiddels zijn ook vele anderen geholpen bij het oplossen van problemen met hun HF versterker.

BACK BACK

Hams

Benelux QRP Club	PGØG techniek	DL1SDQ components
G-QRP Club	DL5JYN QRP stuff	HA8UG amps, components
American QRP Club	ND2X linears HF/VHF	AC6V ant projects, mni info
German QRP Club	KO4NR mods with sweep tubes	DF2OK mni info via links
PA3ESZ mni info	WB6BLD meter scale prog	SM0VPO homebrew

PAØCMU homebrew | NM3E Bird equipment

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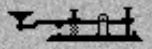
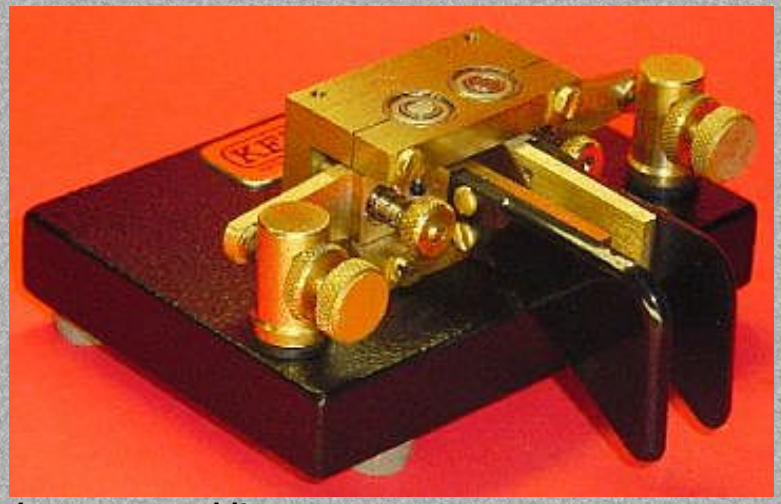
Biografie PAØFRI

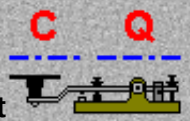
Toen ik 10 jaar was zijn de eerste schreden op het electronicapad gezet met het bouwen van een kristalontvanger. Wie deed dat destijds niet? Niet veel later doneerde een radioknutselaar uit de buurt mij een paar oude "lampen" en andere onderdelen. Daarmee zijn onder zijn supervisie rechtuitonvangers gebouwd.

De volgende stap was de zelfstandige bouw van een supertje met 1.4V batterijbuizen. Destijds een sensatie als je als enige met een draagbare ontvanger (formaat kleine koelbox) op het strand van Castricum en Scheveningen verscheen. De 90 V batterij was veel te snel leeg en een nieuw exemplaar was een behoorlijke aanslag op het zakgeld. Verder zijn veel projecten nagebouwd uit Radio Bulletin, Radio Electronica en Jongensradio delen 1, 2 en 3.

Daadwerkelijke kennismaking met het zendamateurisme vond plaats in 1961. In dat jaar werd ik CW operator van PI1KMA. Één van de verantwoordelijke zendamateurs was wijlen Henk, PAØDB. CW lessen konden op vrijwillige basis gevolgd worden en samen met instructeur van Lieshout heeft hij mij de edele kunst van  morse seinen en opnemen geleerd. Hun geduld werd op de proef gesteld omdat de leerling niet paste bij de gangbare lesmethode, hi! Gelukkig werd de drill losgelaten en met bijlessen in de schaarse vrije tijd is een ontbrekende morsevaardigheid nog betrekkelijk snel opgedaan.



Een paar jaar later is met goed gevolg examen gedaan voor een A-machtiging en de overheid gaf een "bevoegdheid" af. Een zender moest toen een half jaar na het examen ter keuring aangeboden worden als meteen de call werd aangevraagd. Zag men ervan af dan werd een "bevoegdheid tot het bedienen van…" verstrekt. Daarmee is regelmatig als gastoperator bij clubstations of andere zendamateurs gewerkt. De eigen call PAØFRI is pas in maart van 1969 telefonisch aangevraagd en toegewezen. Toen kon dat nog en was het binnen een paar minuten geregeld!



Gestart werd met en EL500 in een eigenbouw 30W CW zender. Ook de ontvanger was een zelfgebouwde Semcoset Semiconda.

Het bouwpakket kwam van Schaart die zijn commerciële activiteiten in zijn woonhuis begon.

De eerste antenne, een 5 m legerspriet, was schuin gemonteerd tegen het 60 m lange metalen balcon. De flat lag op de zesde en hoogste verdieping en daarom was ik tijdens een pile-up er meestal als eerste bij. In 1982/1983 is gewerkt met PAØFRI/OD tijdens een UNIFIL missie in Libanon. Begin 1993 is mij in het voormalige Yugoslavië als UNPROFOR waarnemer in Croatië een machtiging verleend met de call 3A/PAØFRI. Daarmee is in de omgeving van Dubrovnic gewerkt. Vermoedelijk ben ik een van de eersten met een officiële geldige schriftelijke Croatische machtiging, want locale amateurs hadden nog documenten met een YU prefix. Tijdends die periode is ook nog gewerkt op het clubstation van Goràzde in Bosnië.

Verder ben ik ongeveer acht jaar VERON/VRZA QSL mananger geweest voor amateurs in Etten-Leur en twee jaar voorzitter van de VERON afdeling Breda.

De advertentie site www.zendamateur.2dehands.nl is door mij opgestart als service voor collega radiozend-en luisteramateurs. Het is een min of meer zelfstandige dochter van 2dehands.nl. Voor het beheren en bewaken krijg ik geen vergoeding. Helaas word ik regelmatig lastig gevallen met onbeschofte email van lieden die zich niet voor kunnen stellen dat alles pro deo gedaan wordt in mijn toch al schaarse vrije tijd.

Door het experimenteren en af en toe publiceren krijg ik onderdelen van andere amateurs en daarvoor wil ik hen hierbij nog eens bedanken. Hun spullen zijn respectievelijk kapot gegaan, werken nog, zijn ingebouwd of liggen klaar voor een volgend project. Ter geruststelling, ik heb ze nooit verkocht, maar een enkele keer weggegeven aan een andere experimenterende of zelfbouwende zendamateur.

BACK BACK

Links

Zendamateurs pagina	Zendamateurs boogo links	Mods.dk (veel modificaties)
Zendamateurs startkabel	Zendamateur 2dehands	Mirror site voor www.mods.dk
Kent voor componenten		Beoordeling apparatuur (reviews)

Contact

E-mailaddres:	pa0fri@planet.nl.	pa0fri@amsat.org	
Webaddress:	http://home.planet.nl/~fhvgeerligs	http://www.pa0fri.geerligs.com	http://www.qsl.net/pa0fri

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Visitors Bezoeker

@nifty

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This site has moved. Please update your links and bookmarks!

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2004/06/04

These Pages are written and managed by
IDEHARA, Norimichi

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June 4, 2004

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IDEHARA, Norimichi

NOTICE:

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You will be directed to the new site in approximately 5 seconds.
Once there, please bookmark the new site and update any links.

Thanks for visiting! Enjoy your stay in ...

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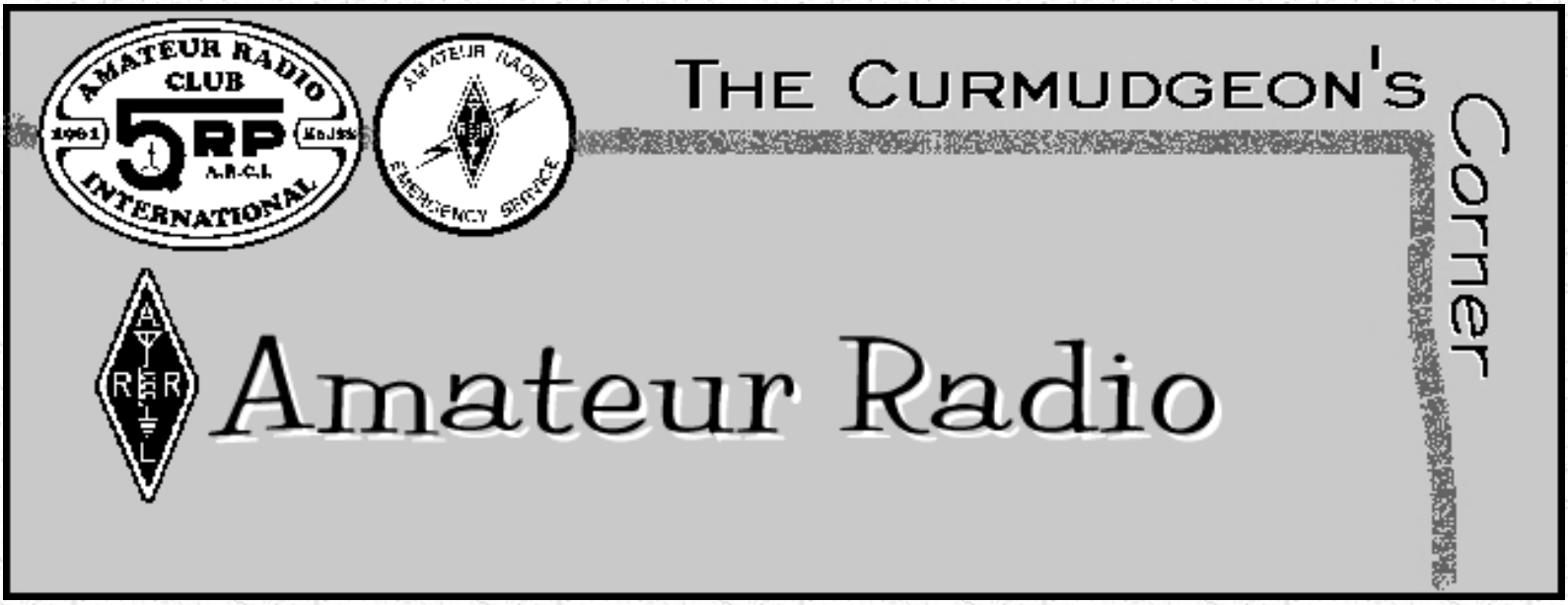
Click HERE for the current FOR SALE, WANTED TO TRADE, and WANTED TO BUY listings!



I can be found, as a friend of mine says "putting little dBs into the RF", as Amateur Radio Operator N5NW. [Click Here to send me email] .

View My Guestbook
Sign My Guestbook

I used to be KM7W. How does a 7-land call end up in the 4th Call District? Simple -- it's a vanity call, and one of the only ones left that contained my initials. I've held four call signs in my time, N4UYT (bad on CW and Sideband), KN4BH (bad on CW!) and KM7W ("you're not from around here, are you boy?" -- HA HA!). Had my old call been better on Morse Code, I would have kept it and not bothered with the vanity stuff.

But, since getting interested in Low Power (QRP) operation, I needed a shorter, more "CW-friendly" call sign.

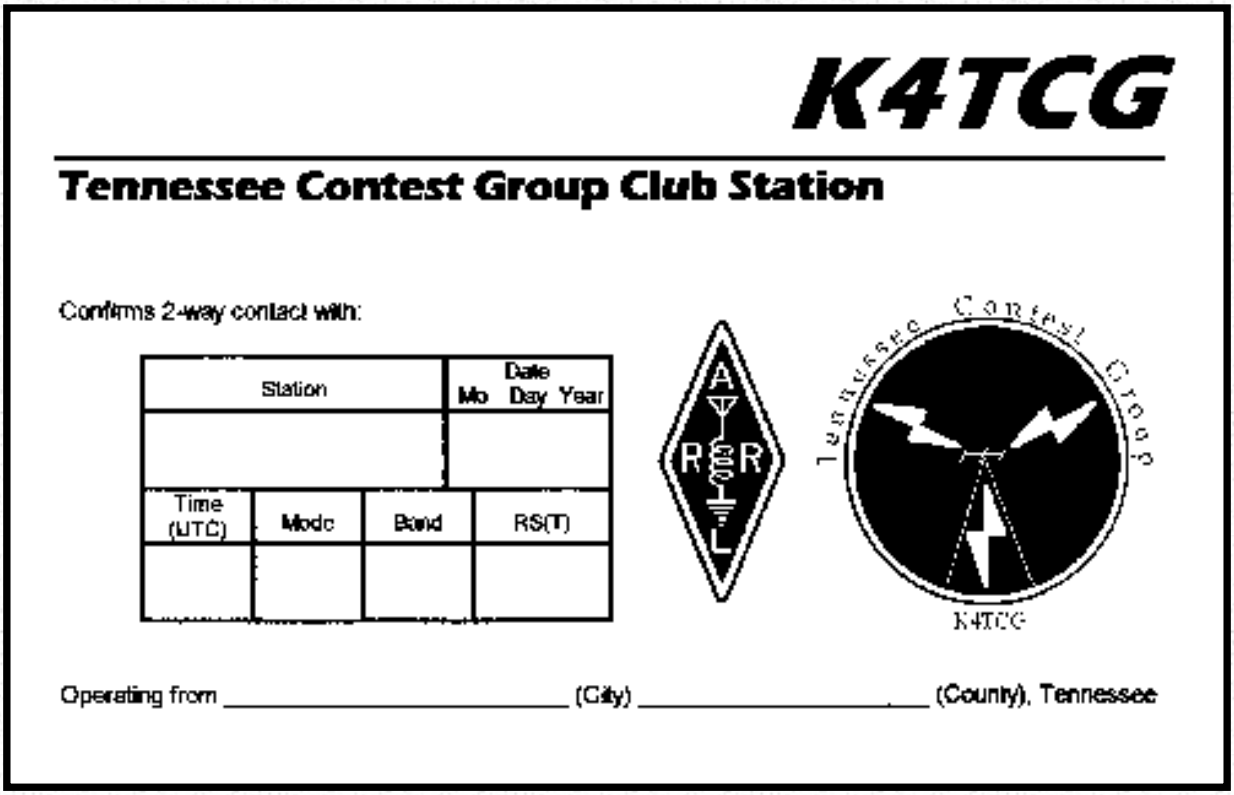
The current station consists of an Icom IC-706MkIIG for VHF and UHF, and a Kenwood TS-570D(G) for HF. Antenna is an off-center fed doublet, vertical dual band (2m/440), a 2m Squalo, and a 6m Horizontal loop. I'm hopeful to get a beam up in a year or so. I have a CMOS III SuperKeyer, Kent paddles, Vibroplex Brass Racer paddles and the Envirotronic 501G paddle through the computer). I've also got a CMOS II SuperKeyer for backup.

Past stations have consisted of homebuilt tuners and power supplies, a Sierra multi-band QRP rig, and a complete Kenwood TS-430s with MARS modifications. Former rigs include a Kenwood TS-570D(G), a couple of Kenwood TS-940sat, and an Icom 706MkIIG. I was the Net Manager (callsign NNNØGBU-3) for Tennessee Navy/Marine Corps MARS. My personal Navy MARS call (officially a NAVMARCORMARS STA) was NNNØRBB. I also have owned a TenTec Argonaut 556 (nicknamed the "Cubbie", because it is a low-power version of the Ten Tec Scout), a homebuilt 38 Special (one of the NorCal Projects), and various VHF rigs, with an MFJ-1278 multi-mode

controller.

The Antenna Farm used to consist of one parallel dipole for the 40 and 30 meter bands. It is made of 450 ohm twinlead, and only requires 100 feet of the stuff for four band construction! E-mail me for details, or check the ARRL Handbook, available from the American Radio Relay League, a national Ham Radio organization. Now, I just have some wire around! You can find more detail on the antenna page. This page also has information on my favorite compact antenna, the Distributed Capacitance Twisted Loop, or DCTL.

I'm a member of the Tennessee Contest Group. I'm currently classified as a pop gun, although I hope to make it up to little pistol some day. My idea for a club QSL card:



The font is Eros Bold ITC @ 34 point for the callsign and 14 point for the text below the line. Other text is in Arial 8 point, except in the confirmation block where 7 point Arial is used.

I'm a member of the QRP Amateur Radio Club International (#7514), the Internet QRP Club (#953), the Alaska QRP Club (#098), and the Northern California QRP Club (#2031). Click on the Club Names to go to their pages -- links for ham radio are abundant elsewhere, and I don't want to bore anyone with another page full of cross-links.

Tennessee Special License Plate application for Amateur Radio. Click on this link. Takes a while to load ... be patient.

The FOR SALE and WANTED TO TRADE/BUY section

I have:

two Heathkit wattmeters/SWR meters (2kW)

Return to The Curmudgeon's Corner (Note: This link leaves this domain to my family page)

[Click Here to send me email] .

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Antenna tech page

New antennas, simulations, antennas related projects and information.

The DDRR antenna
designed by Dr. Boyer for Northrop

The DDRR (Direct driven ring radiator) was invented by Dr. Boyer from Northrop, for military applications in the 50s. It stayed classified until Dr Boyer published an article titled "Hula hoops antennas" in Electronics (Jan 11, 1963). "73 Magazine" published also a two part article from Dr. Boyer titled "Surprising Miniature low band antenna". The article is quite technical and contains a very precise mathematical formulation of the DDRR based on transmission line theory.

In layman's terms a DDRR is just a short vertical monopole (vertical post) attached to a transmission line tuned by a reactance (ring plus vacuum capacitor). The reactance of the capacitor is transformed by the transmission line and will under specific conditions (length, value of the capacitor, etc.) make the vertical post resonant. Slight variations of the capacitor will lengthen or make the transmission line "shorter" and allow tuning of the antenna on a certain range.

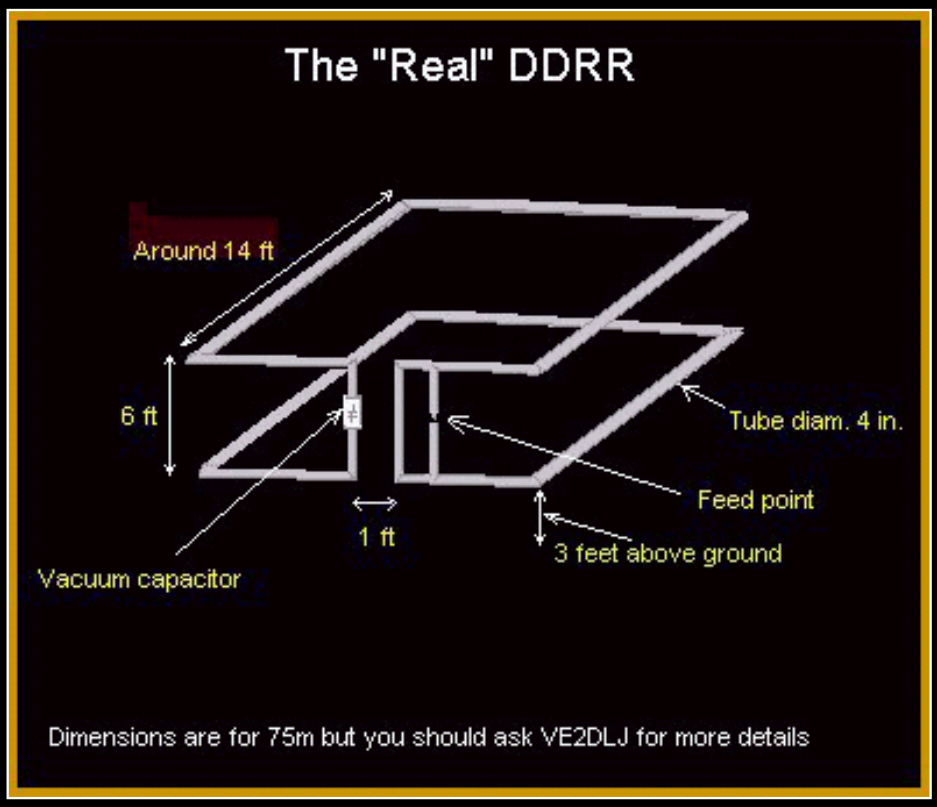
First lets make a difference between the DDRRs patented by Dr Boyer and the ones that we can find here and there in amateur publications. Boyer designed two basic models: the one ring DDRR and the two rings DDRR.

The first one is made of an opened ring made of aluminum tubing (4 in.) over the perfect ground of a metallic structure like a warship, and with two aluminium posts connected at the extremities of the ring. A set of fiberglass post supports the ring. One of the vertical post is attacked at the base by the coax, the other one contains a variable vacuum capacitor to tune the antenna. This faded picture from the 50s from Northrop show a set of concentric DDRR for 2 to 30Mhz. The engineer is standing below the 75m loop (6 feet height) and is looking at a 50Kv vacuum capacitor. This type of DDRR was installed on some special communication and ELINT warships, and was apparently used in the early stages of the Apollo project.

Contrary to most ham publications, Dr Boyer in all the articles he wrote, and in his discussions with VE2DLJ and VE2AMT stressed the importance of using big fat aluminium or copper tubes and to avoid at all cost chicken wire as ground plane. **In a one ring DDRR he insisted on the fact that you needed a solid metallic ground plane**.

The advantages of DDRR compared to a vertical are obvious, especially on a ship. Small size and height, remotely tunable, low noise due to loop structure, extremely low impedance and therefore no or little influence on connected electronic equipment from voltage transients and lightning. Low angle for long range communication and possibility of a secondary high angle for NVIS, indispensable in fleet communications, can be achieved by using a configuration using a central post and two posts with vacuum

capacitors.



The second type is a double ring system isolated from the ground with the same two vertical posts connecting the opened extremities of the rings. One is solid aluminium, the other one contains a Vacuum. A third aluminium post cut in the middle is used to feed the antenna and can be moved to adjust the SWR. The coax enter in the middle of the right post, then up then down to the "feed point" where the shield is connected to the top part and the center of the coax to the bottom part of the post. This configuration creates a coaxial balun for a symmetrical attack. Matching is achieved by moving the "feed point post".

The antenna is standing and supported by very high quality isolating posts made of PVC or fiber glass, and Dr Boyer even recommended beehive isolators at the extremities of the posts to limit losses. The whole structure is placed some 3 feet above the ground.

Contrary to the one ring model, there is no need here for a perfect ground and Dr Boyer was confident that this antenna could perform well even on very poor ground. He referred in his articles to the fact that DDRR are essentially "magnetic" antennas in the near-field and that losses in the ground my magnetic field are negligible. This is not entirely true and computer simulations and experiments by VE2DLJ showed that even if losses in the ground are negligible there is an improvement when radials are inserted.

Tests made by Boyer and his team for Northrop and the US Navy showed that the antenna was very low noise and very low angle. This was done in side to side comparisons made with a collapsible vertical on 160m in the Arizona desert, using a half square mile ground plane made of solid copper sheets soldered together (visible in the picture).

VE2DLJ and VE2AMT found some ten years ago an old copy of the 73 Magazine article and decided to build a prototype. They contacted Dr. Boyer and have numerous discussions with him. He warned them and told them "not to cut corners". His recommendations were:

- Use big tubing: 4 inches or more for 75m.
- Do not use automotive exhaust pipe. It will rust and contacts losses will transform the antenna into a dummy load.
- All contacts and connections must be A1. Solder corners if you make it square, use Penetrox everywhere.
- Use very high voltage vacuum capacitors, and high quality isolators to support the rings.
- No chicken wire. Dr Boyer was horrified by some description of DDRRs using chicken wire. He explained that measurements made by Northrop engineers showed that the near field is in concentric rings on a one

ring DDRR and that chicken wire could add losses.

Our two explorers decided to build the antenna following scrupulously all recommendations. Some years ago, Alex, VE2AMT, made a square DDRR for 75m, with special machined metal brackets in the corners so that it could be disassembled if necessary. He adjusted it and worked two VK and a ZL the first day he used it on the air using 100W. He explained that listening on this antenna was a real experience. He could hear no noise at all and the DX was Q5 all the time, even when other stations couldn't copy the DX. The antenna was at that time sitting on the side of his house on wooden blocks 2 feet above the ground.

VE2DLJ followed suit a few years later and his DDRRs are visible in the Picture page. He is very often in the 75m DX window working VK, ZLs and other pacific stations.

Models

NEC4WIN Model

```
CM    **********    80m DDRR    **********
CM
CM     Designed by JM Boyer for Northrop
CM   Supposed to be Low Noise, DX antenna
CM   VE2DLJ, Tony Cicchetti, is using a full size
CM   version for 80m and can be heard in the DX Window
CM
CM  US Patent #RD26196, RE 3,151,328 Northrop Corporation
CM  "Hula-hoop antennas", Electronics, Jan 11 1963, Boyer.


CE
GND Reference
UNITS Feet
Height 0.88
Over Ground  14  6 (Diel. - Cond. µSiemens)
Circular Boundary
F  3.790
GW  1   3    0.000   0.000   0.000   6.000   0.000   0.000   0.500
GW  2   3    7.000   0.000   0.000   14.200   0.000   0.000   0.500
GW  3   5    14.200   0.000   0.000   14.200   14.200   0.000   0.500
GW  4   5    14.200   14.200   0.000   0.000   14.200   0.000   0.500
GW  5   5    0.000   14.200   0.000   0.000   0.000   0.000   0.500
GW  6   3    0.000   0.000   6.000   6.000   0.000   6.000   0.500
GW  7   3    7.000   0.000   6.000   14.200   0.000   6.000   0.500
GW  8   5    14.200   0.000   6.000   14.200   14.200   6.000   0.500
GW  9   5    14.200   14.200   6.000   0.000   14.200   6.000   0.500
GW  10   5    0.000   14.200   6.000   0.000   0.000   6.000   0.500
GW  11   4    6.000   0.000   0.000   6.000   0.000   6.000   0.500
GW  12   4    7.000   0.000   0.000   7.000   0.000   6.000   0.500
S    1  48    100    0
LC   1  43    0    41.8
Coax  1
```

NEC4WIN95 Beta v1.00.5 Log File

DDRR.N4W
Antenna Height is : 0.26 meters
Ground Diel. = 14 Cond. = 6 µSiemens
Frequency = 3.790 Mhz
Wave Length = 79.103 Meters
Load # 1 = 0.000E+00 +j -1.005E+03 at Pulse 43

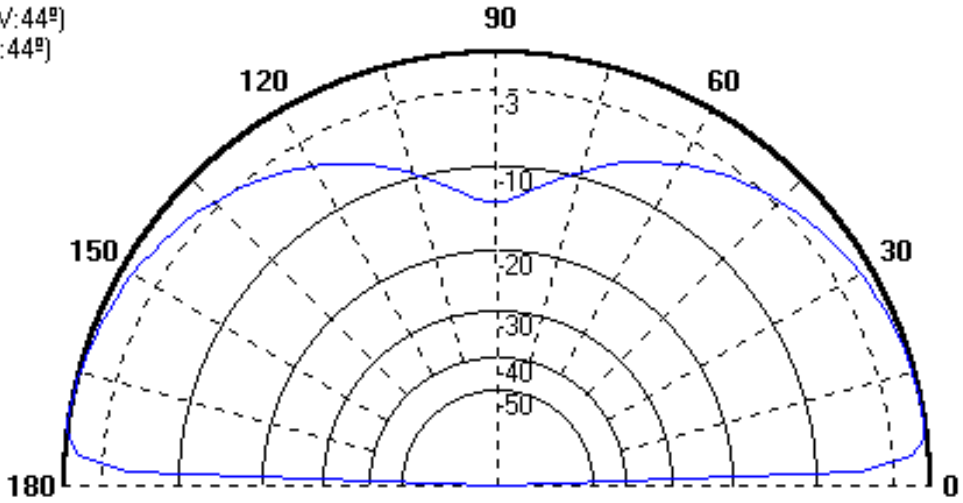
Impedance = 0.46 - j 0.08 Ohms at Source 1
SWR = 2.19 with 1 Ohm Coax

DDRR.N4W Zenith
Antenna Height is : 0.26 meters
Ground Diel. = 14 Cond. = 6
Z1 = 0.46 - j 0.08 (2.19)
Height = 0.268 m
Max = 1.13 dBi **<----- NO RADIALS**
Lobe at : 152º (BW:48º)
Lobe at : 28º (BW:50º)

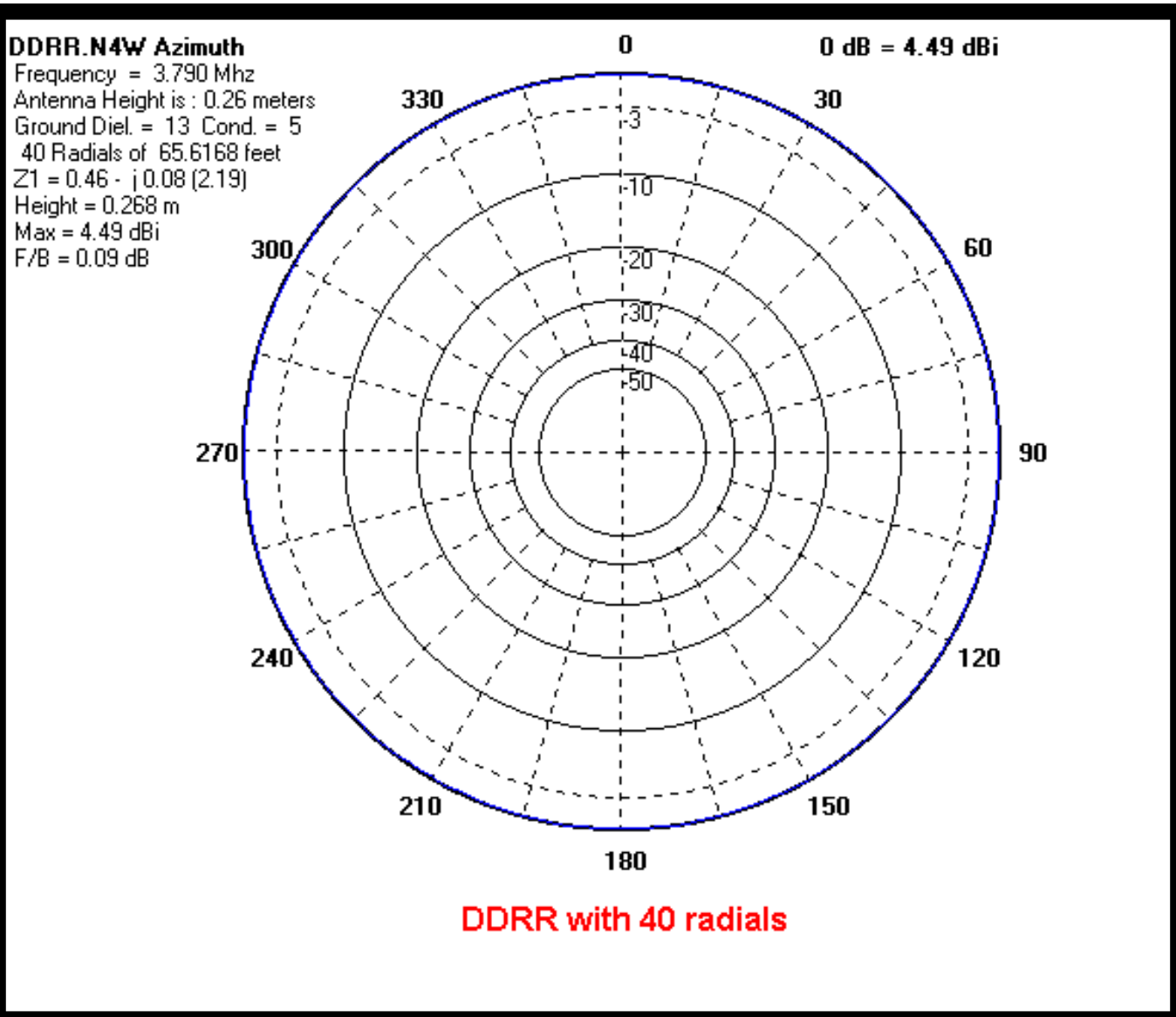
DDRR.N4W Zenith
Antenna Height is : 0.26 meters
Ground Diel. = 13 Cond. = 5
40 Radials of 65.6168 feet
Z1 = 0.46 - j 0.08 (2.19)
Height = 0.268 m
Max = 4.47 dBi **<--- 40 radials of 66 feet**
Lobe at : 170º (BW:44º)
Lobe at : 10º (BW:44º)

DDRR.N4W Zenith **0 dB = 4.47 dBi**

Frequency = 3.790 Mhz
Antenna Height is : 0.26 meters
Ground Diel. = 13 Cond. = 5
40 Radials of 65.6168 feet
Z1 = 0.46 - j 0.08 (2.19)
Height = 0.268 m
Max = 4.47 dBi
Lobe at : 170º (BW:44º)
Lobe at : 10º (BW:44º)



DDRR with 40 radials



AO Model

DDRR for 80m
Ground 13 5

1 zone 50 radials #12
13 5 0 100

3.795 MHz
10 aluminium wire, feet

```
10    0   0   3   14  0   3   0.3
10    14  0   3   14  14  3   0.300
10    14  14  3   0   14  3   0.300
10    0   14  3   0   1   3   0.300
10    0   0   9   14  0   9   0.300
10    14  0   9   14  14  9   0.300
10    14  14  9   0   14  9   0.300
10    0   14  9   0   1   9   0.300
10    0   0   3   0   0   9   0.300
10    0   1   3   0   1   9   0.300
```

1 source
wire 9, center

1 load
wire 10, center 10 pF

ANALYSIS

Impedance 0.678 - j 18.7 Ohms
SWR 84.01

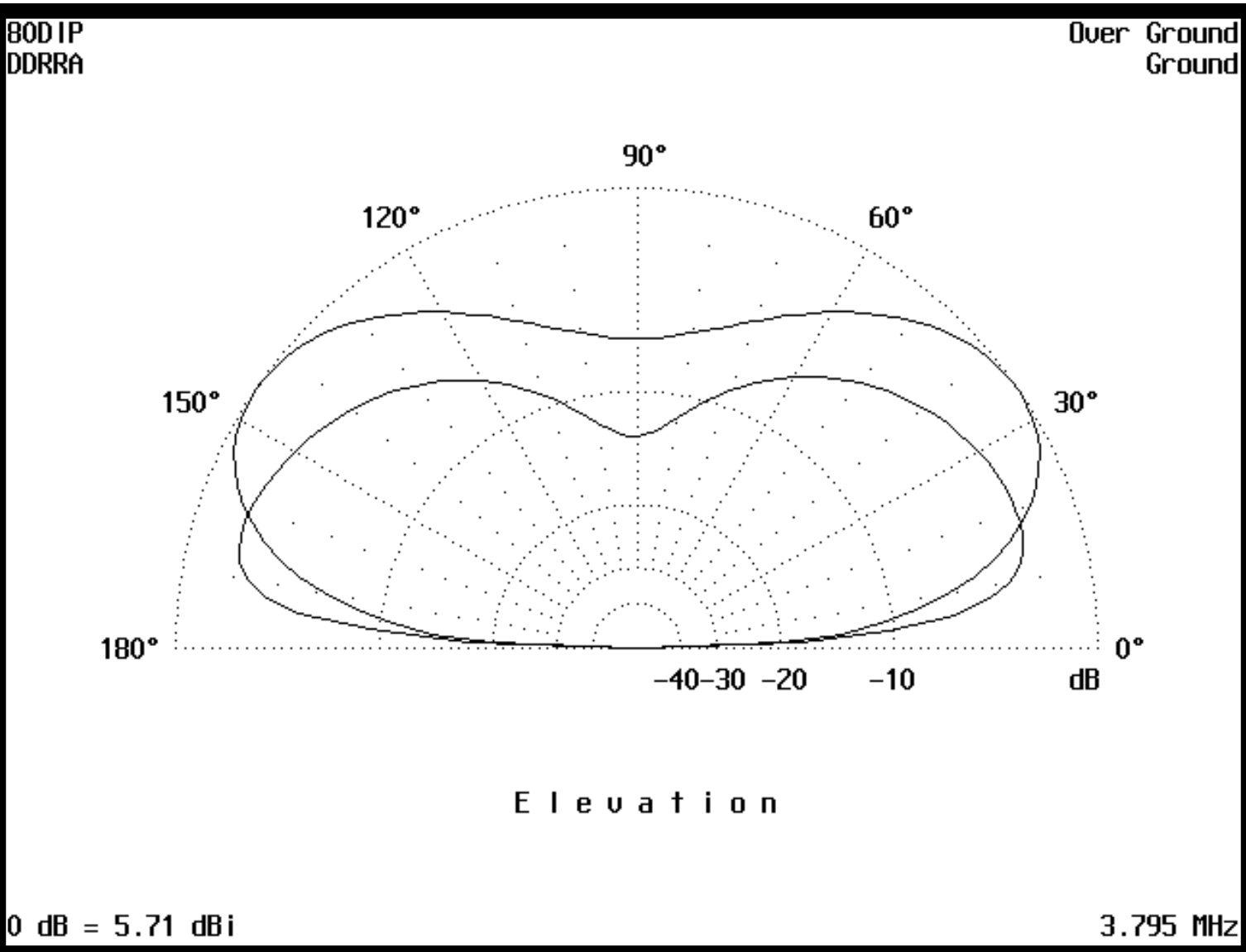
Wire Losses 0.33 dB
Efficiency 92.6%

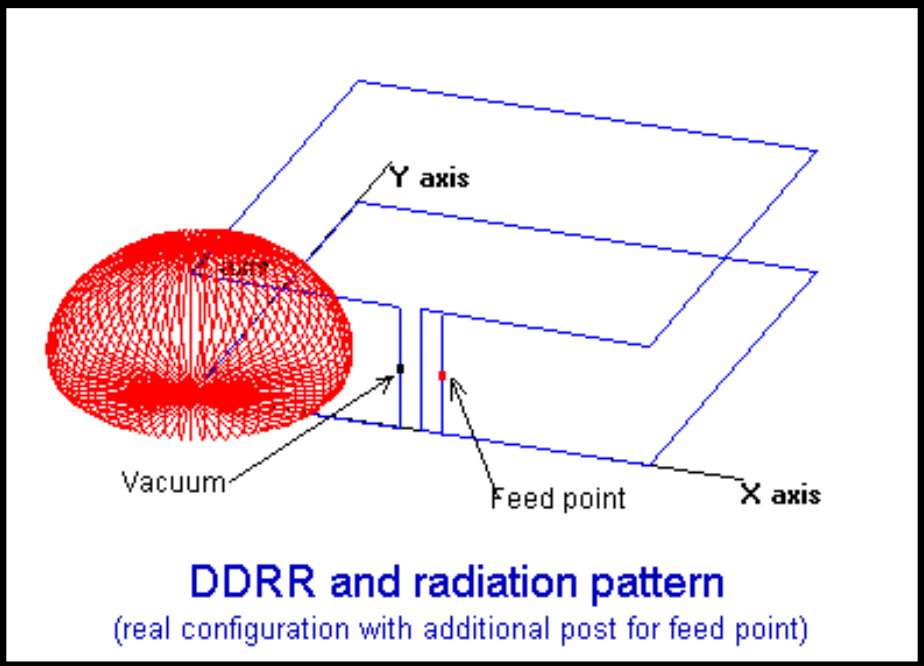
At 18.0ø Elevation:

Forward Gain 3.41 dBi
F/B -0.34 dB

Maximum Gain 3.81 dBi at 132ø Azimuth
Radiation Peak 164ø Elevation (3.77 dBi gain)

DDRR with radials vs a Dipole at 130 feet DDRR is better at low angle (<18 deg)





DDRR and radiation pattern
(real configuration with additional post for feed point)

Comments and remarks

The DDRR impedance is very very low. At 0.5 ohms, any wire or connection losses will transform the antenna into a dummy load. This is why one needs big fat tubes. Losses in the tuning capacitor must be minimized too and a vacuum of very high voltage is indispensable.

Contrary to Dr Boyer's predictions, **the addition of radials increased gain and lowered radiation angle**. This was confirmed by VE2DLJ experiment's. We see from the zenith patterns that the TOA is below 20 degrees and that the antenna is a low angle radiator. **This could be OK for long range DX like VK and ZLs but it is not very good for medium range DX**. This was confirmed by long term testing over a year and a half period. Tony, VE2DLJ, has a horizontal loop "reference antenna" and the loop will always beat the DDRR toward Europe in transmission even in Winter. The horizontal loop will also give better reports (in average) than DDRR on long range DX *outside of Winter DX period* but starting around beginning of October the DDRR will start beating on average, the horizontal loop by one to two S units in transmission on VK, ZLs and JAs.

In reception, the DDRR (according to Tony, and most hams who listened on them) will always beat the loop for QRN and man made QRM. Signals will be sometimes one to two S units lower but always perfectly Q5. The bandwidth is some 20 to 30Khz and the Q is in the 300 to 400 range, so the antenna is very selective which is a big advantage.

We can see from the comparison of a **dipole at 130 feet vs the 3 feet high** DDRR that the DDRR is better on low angle signals. The difference is even more important for a lower dipole.

What are the advantages of the DDRR? Low noise antenna, low angle. "Small size" requiring little real estate. Could be easily installed in a backyard at ground level or on a flat roof. No tower needed, no climbing. Immune to noise, static electricity and ligthning. Tunable, very high Q. Boy that seems great!

Does that mean that I will "recommend" a DDRR? Hell No! To build a good system you need some 100 feet of 4 in. tubing, then solder the corners (or use special machined brackets) to reduce losses. If you want to transmit full legal power you need 50 KV vacuums, and they are not cheap. Add the cost of good isolators to support the rings etc and you end up with a two to three thousand dollars antenna. Even if ground does not influence much the DDRR in the near field, ground is important for the first reflection and you need good ground around the antenna. If you have neighbours the low angle may create TVI. Another major point is directivity. The DDRR doughnut is omnidirectional and to work DX you need directional antennas. That's the only way to eliminate QRM and concentrate the energy in one direction. This means that one DDRR is not enough and you will need at least two.

To build low band DDRRs, there are cheaper alternatives using salvaged Heliax. Three or four VE2 designed DDRRs using plumbing copper tube, wire and Heliax with variable success. The only area where price/performance is interesting, is in VHF/UHF. At these frequencies, DDRRs are easy to build and very small. Otherwise, stay away from them unless you know what you are doing.

Madjid
VE2GMI

Patents: US Patent (J.M. Boyer) #3,151,329; #3,247,715; RD26196 all assigned to Northrop Corporation

Note: Tony, VE2DLJ can be found in the 75 DX window on most mornings, sometimes evenings at sunrise and sunset.

DDRR2 Article

If you have questions, comments, critics, suggestions click here: art@orionmicro.com
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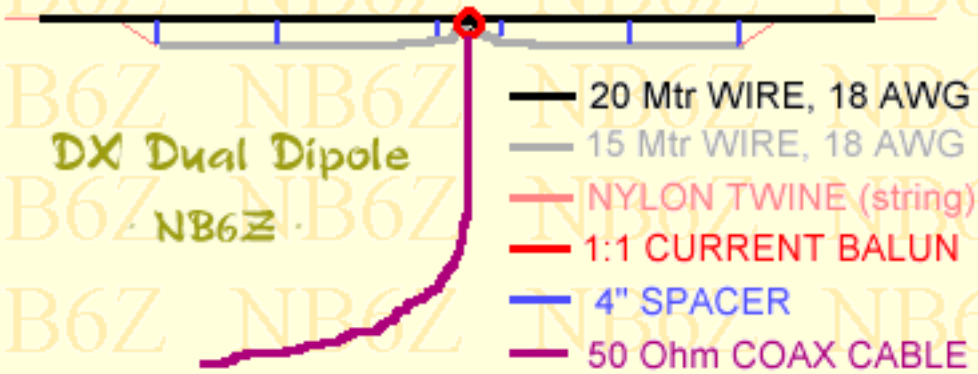
First the Dipole...

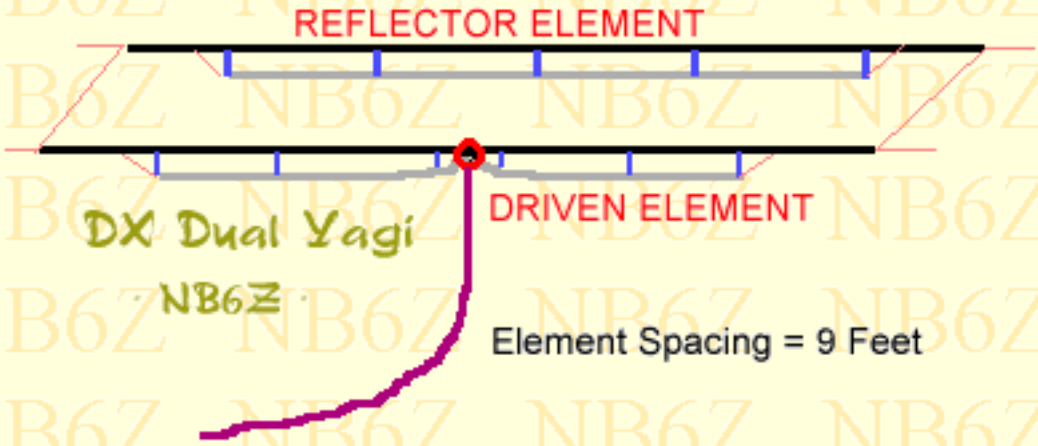


The antenna diagram you see before you is nothing more than two dipoles fed from the same coaxial line. I call it a " DX Dual Dipole" because it is cut for two very useful DXing ham bands. There is no harmonic relationship between 20 and 15 meters, and so they "play" well together when sharing the same transmission line. The wire length cutting chart for HF ham band dipoles is found in any antenna handbook. (The half wave dipole cutting formula is: Length in feet = 468 divided by the Frequency in Mhz.) Both dipoles are cut per the chart (or formula) to the low end of each band, leaving a little extra wire length for SWR tuning. Construct the dual dipole using the techniques described for the NB6Zep Antenna and make the wire spacers from plastic coat hangers as described there. Follow the instructions in the handbooks to make a 1:1 current type balun device using a small toroid core. (The smaller T130 red HF material cores work fine for several hundred watts.) The use of a balun is not strictly necessary for the dual dipole, but it is strongly advised if you are going to add a parasitic element (reflector) as described later. If your feed point is not supported, as was the case for me, you will want to use a strong stranded wire (#20 thru #12 AWG) and light weight coax.

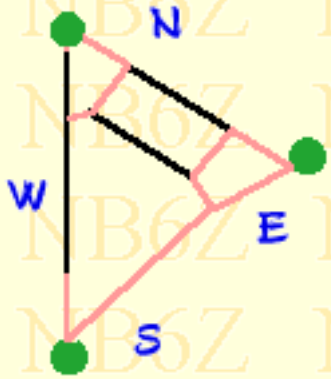
Here is a **HOT TIP** for light weight and portable operation. Use short runs of RG-174 coax for low power operation at HF frequencies. I tested 30 feet of this very small coax at 50 watts before beginning construction and found no objectionable loss or leakage at 14 Mhz. (Do not use this coax at power levels greater than 100 watts or with antennas that require a tuner!) Another advantage with RG-174 is you can make a current balun (coax choke) simply by wrapping 10 turns of the coax around a T150 toroid core at the end of the cable where you attach it to the wires. I have been using this cable successfully for several years...

To tune the antenna, each dipole must be trimmed separately by cutting small equal length pieces from each end until a low SWR (below 1.5 to 1) is achieved for the low end of the band. (You may tune them to the desired frequency if you do not intend to add a parasitic element later.) This can be done with the antenna at six feet off the ground, but expect the SWR to change slightly when you bring the antenna to full elevation. (Always listen for a clear frequency before testing SWR. SWR tuning is not needed for receive only operation.) There will be interaction between the two dipoles, so alternate between dipoles until you find the right match. (Note: Yes you can add additional half wave elements for other HF bands to this configuration. Adjusting multiple elements for a low SWR becomes problematic and will take some patience. I recommend a multi band wire with tuner such as the NB6Zep.)

...then the Yagi.



The two element yagi style antenna I am most comfortable with is one that employs a reflector instead of a director element. The yagi I chose to build has a parasitic reflector nine feet behind each dipole. (Nine feet was a compromise for dual band operation.) My computer modeling looked good with standard reflector lengths, so I cut the reflector wire lengths to what I found in the antenna handbooks for 2 element yagis. You should have reasonable results if you cut the wires to 5% greater than the dipole length (dipole x 1.05) for each band. When the reflector is in place behind the driven element as shown, you should notice a slight change in the SWR. This is due to mutual inductance as the reflector is inducing power from the driven element. The effect is to lower the impedance of the driven element. If you trimmed the dipole SWR for the low end of the band as previously instructed, you may find the resonance is now in the middle of the band. However, you may need to do some final trimming on both driven elements to compensate and make a better match at the transmitter side of the coax. When you are satisfied with the SWR on both bands, raise your yagi as high as possible and free from large objects and other tuned antennas.



I have my dual yagi at 30 feet above the back yard and pointed at 33 degrees towards Western Europe from Beaverton, Oregon. A major challenge was to figure a way to support the yagi in the direction I wanted it to operate. I used two trees supporting an existing 15 meter EDZep wire and used a third tree to triangulate enough nylon twine to create the configuration shown. My Dual Band Yagi has been in place, with zero maintenance needed, for over two years... A ***closer view*** shows more detail of the wire spacers, (note the NB6Zep at 40 feet).

Basic Performance: The antenna modeling was done with all the wires in the yard included in the model to get the whole story of how well this antenna should do for me. First on 20 meters... ***CLICK for DETAIL***. Slightly better results is predicted for 15 meters... ***CLICK for DETAIL***. Needless to say, after seeing these predictions, I was excited about building and testing this antenna! Comparison testing has shown that the gain from the yagi is close to what the model predicted. Granted, it is impossible to measure a 3.6 dB gain while operating under skip conditions, but the yagi out performed all wires I tested it against. The Front to Back ratio is a bit hard to confirm from

Oregon; but I note that my 15 meter EDZep used on 20 meters always works better to the South Pacific than does the yagi. The differences between the yagi and the NB6Zep (SE and NW) on 20 meters is typically greater than 20 dB for skip conditions. The beam width is noticeably broader on the yagi than on the NB6Zep and I can use it for contacts to the NE sections of the USA.

With the yagi antenna I am able to work any European station I hear on 20/15 meter CW, MFSK16 or Hellschreiber mode with 50 watts. I can hear good PSK31 sigs from over the pole, but still have a hard time decoding them here in the North West.

Would you rather use a tuner instead of manually trimming each dipole? The Super Zep antenna is a more "elegant" solution for a multi-band wire beam.

"Push-button Menu of Stuff"

Rotatable, Self Supporting Dipoles

KQ6RH

(C) 1998, 1999, 2000

Ray Jurgens

(Up-Dated 2/25/2000)

Rotatable Dipoles:

There are quite a few ways to make full-sized dipoles that can be rotated. How you attack this mostly depends upon what band is to be covered (basically how big) and whether the wire is to be horizontal or is permitted to slope downward from a central post. In the horizontal case, the wire is threaded through the spreaders and may extend out the ends. The ¼" tubes have an adequate ID that #16 wire is easily passed through the tube, so extenders can be added to the usual fiberglass ½" tubes that fit the hub. In doing this, you should be aware that the velocity factor will be less than unity, so the physical size of the dipole will be slightly smaller than that of free space. In order to make connections to the feed line at the hub, two ¼" diameter holes have to be drilled at an outward slant into two opposite spreader sockets. These should be drilled at about a 45 degree slant beginning about ½" out from the center. A ten meter dipole requires no extenders. Longer wavelengths require extenders, and the 20 meter dipole may require wire extending slightly beyond the extended spreaders. In general, feed the wire through the extender first, then into the ½" tubing, then slide the extender into the half inch tubing and push the wire beyond the end of the spreader about 4". Feed the wire into the hub and up through the access holes that you drilled. Then push the spreader into the hub. Now, from the tip of the spreader pull the wire until there remains just enough wire at the hub to make the connections to the feeder. Adjust the length of the spreader extenders, and tighten the hose clamps. Leave about a foot of extra wire beyond the extender. You will then need to trim this to get proper resonance once the structure is in the air. In the case of ten meters, you are done, simply mount the hub on the mast and put it up. No guy lines are needed if you don't mind a bit of droop. In the case of 20 meters, there is much more to do. Here, the length of the spreader will be about 15' if you have a 1' overlap with the extender. So you will need a central guy post 6' long, i.e., use a full 8' section of 1" tubing with 2' below the hub. You will need guys to both the inner spreader at 8' and the outer extender at 15' up to the central hub for both spreaders. You also may need rotational stability if you want this to settle down after rotation or gusts of wind. The two unused socket holes are there for a reason, so, fit two 6' or 8' (if you have room) ½" spreaders in these sockets and guy them in the same manner as before at the 8' and 15' locations. Always set the guys from the inside first, then add the outer ones. This is still a fairly

loose structure since only gravity holds it in the downward direction. If this structure is still not stiff enough, you can guy downward to the mast as well. The limiting tension is set by the point where the extenders begin to buckle. That turns out not to be a whole lot of tension, because a 7' section of the ¼" tubing sets the limit.

The second procedure is to make an inverted V dipole, where the antenna is the upper guy lines from the center pole out to the spreader tips. For the ten meter case, this is nearly identical to the ground/counterpoise discussed in the Quick Vertical section. In that case, there are no extenders, so the construction is very simple. In the 20 meter case, all the same problems are encountered as above except that the wire load is acting as the upper guy lines rather than being in the spreaders. We also suggest using light wire for the 20 meter version. In fact, # 18 or # 20 hook-up wire works well, and the insulation should be left on. We prefer the un-tinned type that is commonly available at Radio Shack. Using, the 6' center pole, the length to the tip should be just about correct, however, the insulation slightly reduces the velocity factor, so you can shorten the extender or use a small length of Nylon fish line to extend the wire.

Note, the 17 and 20 meter versions of these dipoles are fairly large structures and can not be built up in small spaces. They are also rather flimsy, and go through lots of distortion when being tipped up. These are better erected from a push-up mast with the rotator near the top of the mast. This Dipole antenna gives the same gain as all other dipoles, however, the Half Square is a much better DX antenna for a given elevation and may be worth the extra effort. All parts used for the construction of the dipole can be used to construct the Half Square, so there is no loss in investment if you decide to switch. Note that 6 and 10 meters require no extenders, but we do recommend that you use guys from the tips to the center post. The post should extend about 3' above the hub.

Parts required for all 6 and 10 meters versions:

Item	Quantity	Description
1	1	HUB 4-050-100, Central Quad Hub (RFJ)
2	2	8' 1/2" OD fiberglass tubing, spreaders (MGS)
3	1	8' 1" OD fiberglass tubing, boom (MGS)
4	2	GT 4-050 1/2" Guy Ties, for tips of spreaders (RFJ)
5	2	GT 4-100 1" Guy Ties, for tips of boom (RFJ)

Extenders are required for 12 , 15, and 20 meters. To determine how much 1/4" fiberglass to buy, you need to calculate the approximate length required for the dipole. If the wire is to be inside the fiberglass, the velocity factor is slightly less than 90%. The size of a typical dipole is given approximately by 468/fMHz. This formula has a small correction factor for finite wire diameter and end effects. When the wire is inside the fiberglass tubing, the appropriate factor is about 435/fMHz, so the lengths of the spreaders require for 12, 15, 17, and 20 meters is roughly as 8.72', 10.25', 12.02', and 15.33'. Assuming 6" overlap and 8' lengths of 1/2" OD spreaders, the extenders will have to be 1.22', 2.75', 4.52', and 7.83'. Obviously, there is no compelling reason to cut the 8' of 1/4" OD tubing for the 20 meter spreaders. You can get both a 15 and 17 meter extender our of a single 8' length of tubing. 15, 17 and 20 meters require lateral guys to increase the stability. This requires two 4' lengths of 1/2" OD tubing inserted in the two remaining sockets. Guys should be run to the tips of both the 8' dipole tips and the extended tips. This is also true from the central Guy post. The guys can be either 50 lb. test Nylon fishing line or Kevlar thread. The photos associated with the Half Square antenna show structures built with both fiberglass and PVC.

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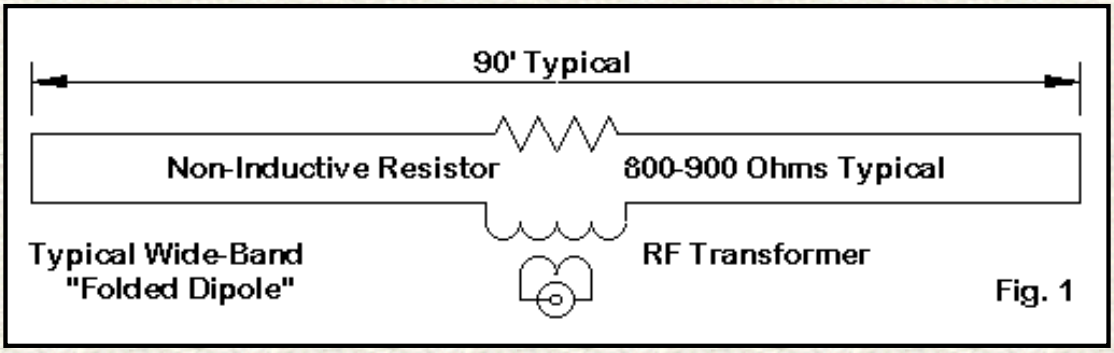
Notes on the Terminated Wide-Band "Folded Dipole"

L. B. Cebik, W4RNL

As space for antennas continues to shrink in the present era of smaller urban and suburban yard, hams have begun to turn to 1-antenna solutions to their operating needs. Among the choices for a horizontal antenna that operates on all of the HF amateur bands, the wide-band "folded dipole" (WBFD) has been gaining popularity. I thought that it might be useful to do some comparative studies using this antenna as a base-line.

The basic WBFD looks something like **Fig. 1**.



The antenna design appears to be a folded dipole. However, a folded dipole is a resonant antenna, while the WBFD is designed to operate with a low feedpoint impedance across a wide range of frequencies. Moreover, the WBFD contains a non-inductive terminating resistor usually located at the point in the loop directly opposite the feedpoint. Normally, the resistor is in the 800-900 Ohm range. This impedance is roughly replicated at the feedpoint. Therefore, builders install a 16:1 RF transformer (either of transmission-line transformer or normal transformer design) at the feedpoint. The result is a low SWR value for 50-Ohm coaxial cable across the entire frequency range.

For receiving use, such as in SWL service, the terminating resistor can be a low wattage carbon type. For transmitting service, the resistor must have a power value capable of dissipating a fair share of the applied power. The exact amount will vary with frequency, but commercial versions of the antenna are often rated for reduced power at the low end of the operating range, where power dissipation is highest.

Commonly, WBFD antennas are offered in a 90 to 100 foot length (27-28 meters) for service between 2 and 30 MHz. However, one can build WBFD antennas in almost any length. Only the effective operating range of frequencies will change.

Since we may also construct doublets of the same length and feed them with parallel transmission line to an antenna tuner, it seemed fair to compare the gain of such a doublet with that of a WBFD of the same length across the 2-30 MHz range. The model I chose for the WBFD is 27.2 m (89.25') long, with the wires separated 0.2 m (7.8"). The terminating resistor is 820 Ohms, a standard value used in some commercial models. The wire is #14 AWG. The doublet is a simple length of #14 copper wire exactly as long as the WBFD.

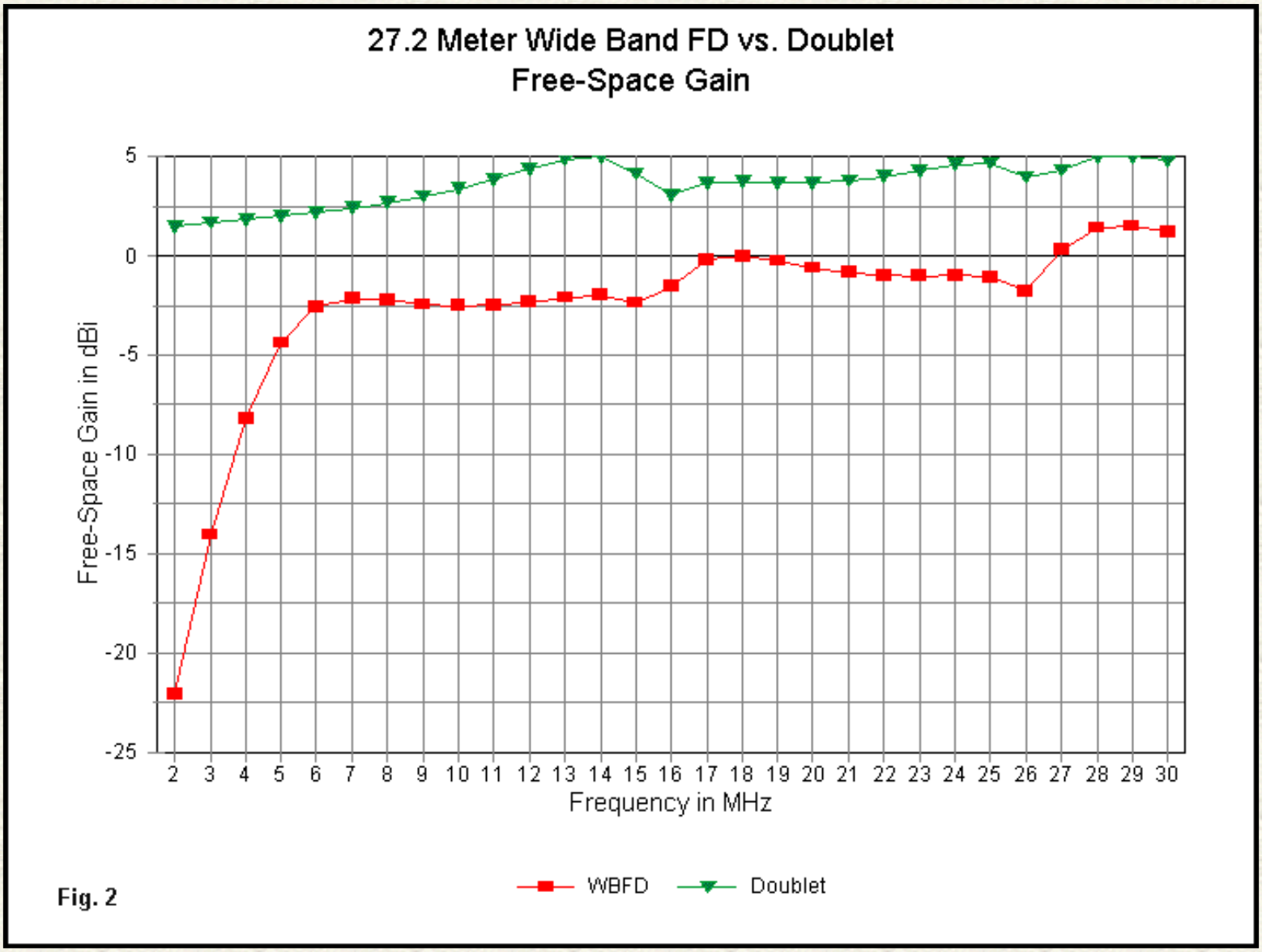


Fig. 2

Fig. 2 compares the free-space gain of the two antennas at 1 MHz intervals from 2 to 30 MHz. Since the elevation angle of maximum radiation will be the same for both antennas for any height above ground and for any ground conditions, any differences that show up in the free-space model will also show up in actual antennas at any height above ground.

Several instructive notes emerge from the comparison of gain in **Fig. 2**. First, the overall average difference in gain between the two antennas is nearly 6.3 dB, with the advantage going to the doublet. If we neglect frequencies below 7 MHz, the average difference diminishes to 5.0 dB. For most of the range of use of the WBFD, then, there is about a 1 S-unit deficit in gain relative to a standard doublet of the same length in the same position.

Second, the WBFD gain curve displays a significant knee--a frequency below which its gain deteriorates rapidly. In the case of the current model, that frequency is about 6 MHz. At or below the knee-frequency, the terminating resistor dissipates more and more of the power. The result is not only a large decrease in gain and higher temperature stresses on the resistor, but as well, very low SWR values at the feedpoint. The knee-effect as the WBFD becomes significantly short relative to the length of a resonant dipole easily accounts for the need to derate the antenna relative to transmitting power below a certain frequency.

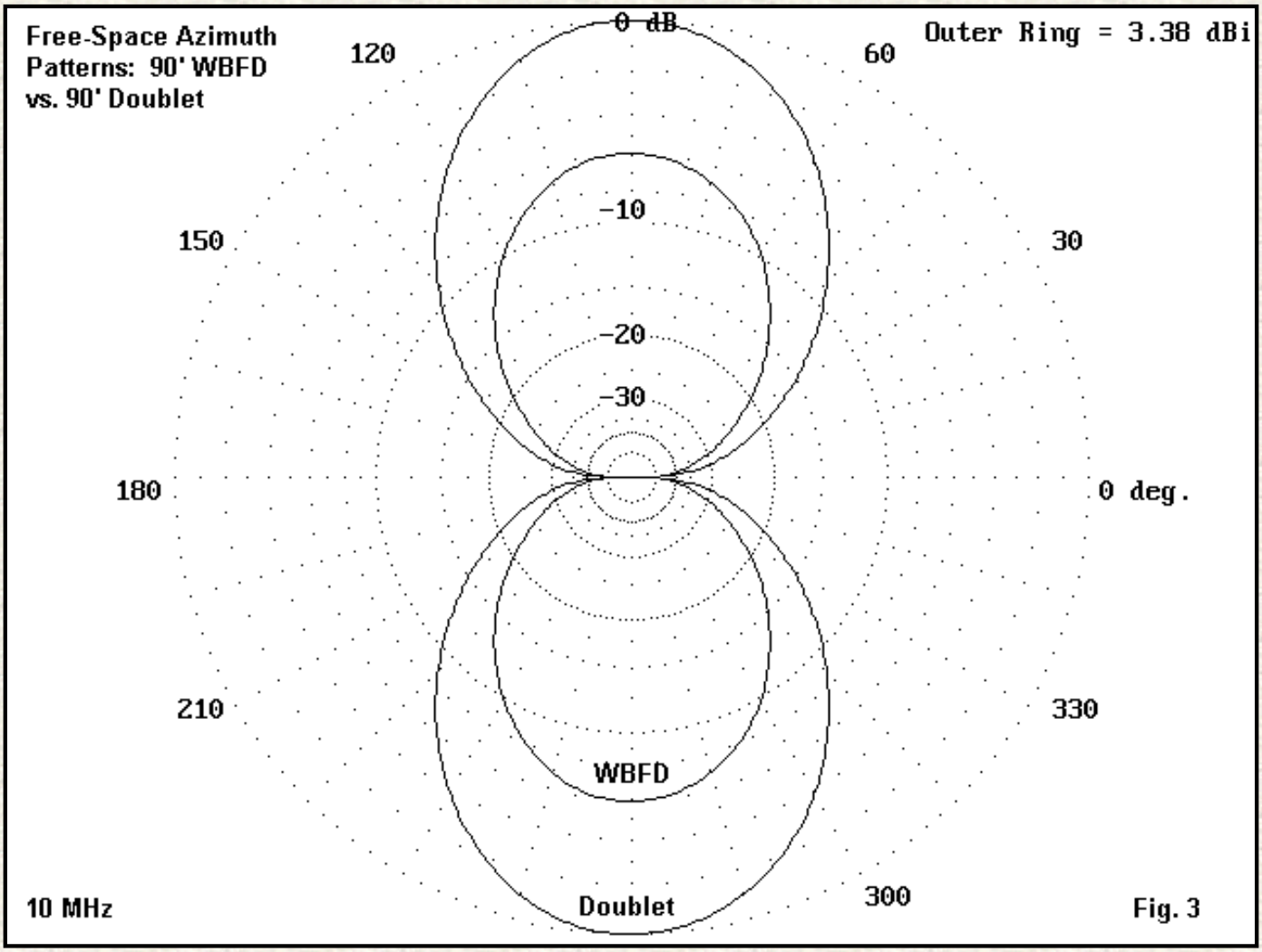
The deficit in gain is not necessarily a disadvantage for receiving purposes. Modern receivers tend to be equipped with receiving pre-amplifiers that the user can switch in as desired. The gain may range from 10 to 20 dB, depending upon design, and in some receivers may be stepped or variable. Therefore the gain deficit can be largely made up in the

lower HF range. Moreover, the basic receiver, apart from preamplification, already has excess gain that is rarely used in the lower HF region.

In addition, one of the major problems in reception in the lower HF range, especially with respect to SW broadcast stations, is front-end overload from excessive signal strength. The overload also tends to produce spurious products within the receiver. Hence, reduced gain of the antenna can be in some circumstances an advantage rather than a disadvantage. Combined with the RF attenuator built into many receivers--which may be a single reduction value or stepped--the WBFD offers a potential for excellent lower HF reception, free of some of the problems that occur with higher gain antennas.

Because the WBFD is also a closed loop with a terminating resistor, many users claim quieter reception relative to doublets for a given receiver input signal strength. The degree to which this is both true and separable from the freedom from front-end overload is difficult to determine. Nonetheless, SWLs have found the WBFD a very useful tool for their efforts.

In order to establish that the WBFD has the same pattern as a doublet of the same length for any given frequency and height above ground, let's look at a couple of sample free-space patterns. For example, see **Fig. 3**.



The 27.2 meter WBFD and its comparison doublet exhibit a bi-directional pattern at 10 MHz. The shape of the pattern is identical, with only the 6 dB gain differential separating the two antennas. The -3 dB beamwidth points are also virtually identical. Since the take-off angle (elevation angle of maximum radiation), the reflection from a given set of ground conditions, and other such factors are not dependent upon signal strength, the two antennas would also show elevation patterns for any equal antenna height that are likewise congruent.

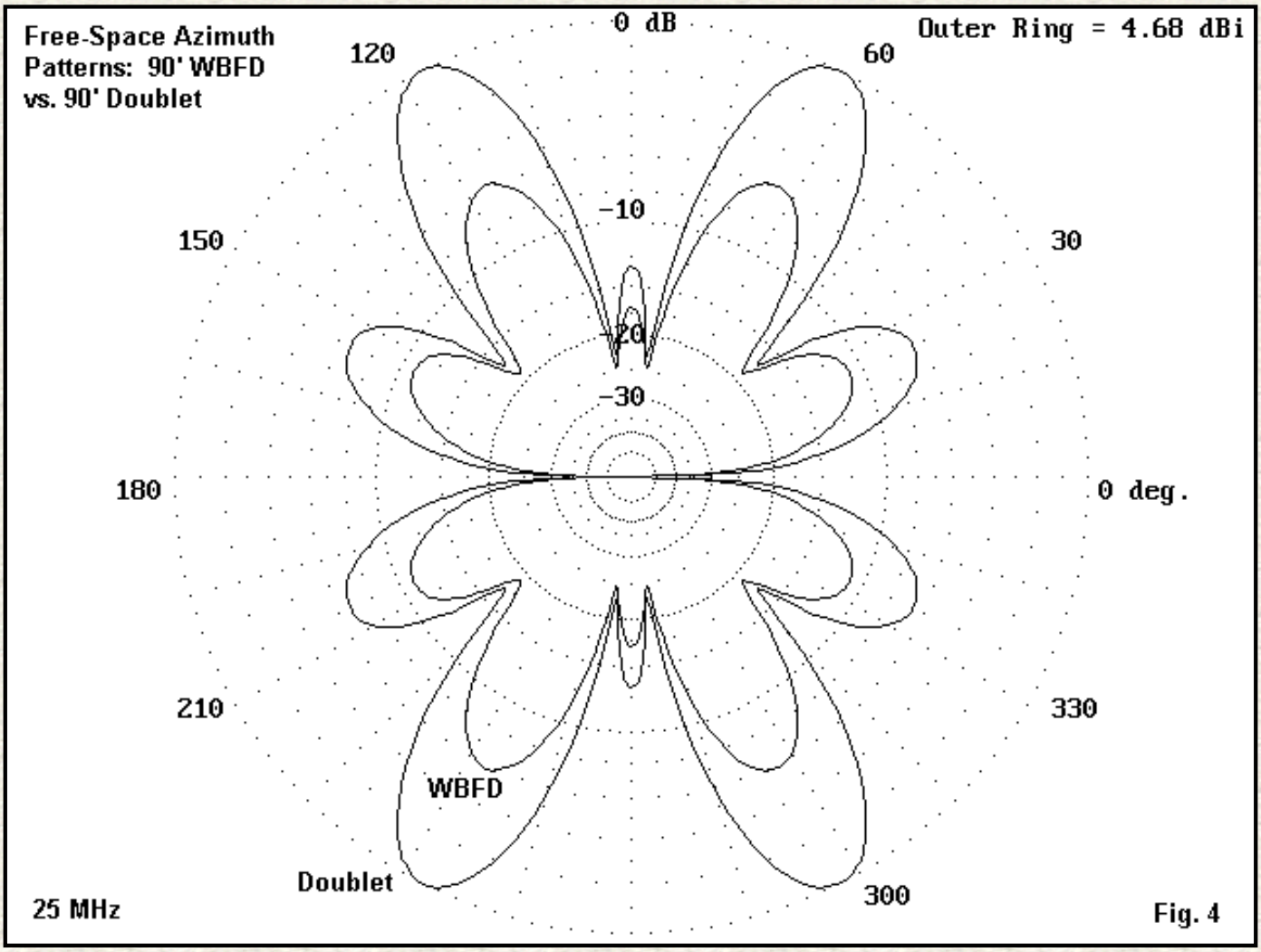


Fig. 4 shows comparative free-space azimuth patterns for the two antennas at 25 MHz. The WBFD pattern is simply a "mini" version of the doublet pattern, with about a 6 dB difference in strength.

There is an additional point in displaying these patterns. The exact pattern of lobes and nulls in the azimuth readings for a WBFD is identical to that of a doublet. As the length of the antenna exceeds 1.25 wl and approaches 1.5 wl, the bi-directional pattern at lower frequencies will break up into a collection of lobes and a collection of nulls. Therefore, the antenna is variably selective in its favored directions of good signal strength as one changes frequency. Those who contemplate installing either a doublet or a WBFD antenna need to consider well the patterns at key frequencies of interest in order to orient the antenna for maximum effectiveness.

The antenna type has also been used vertically to provide omnidirectional coverage. However, in this orientation, when the antenna exceeds 1.25 wl in over length, the pattern begins to show primarily high angle radiation--exactly the opposite of what one normally desires from the upper HF band. As a result, some installations may use a pair of vertical WBFDs for full low-angle HF coverage.

A Note on Knees and Length

The knee we observed in the gain of the 27.2 meter WBFD is interesting, since it suggests that we may vary the low frequency gain by changing the length of the antenna. Changing the length, of course, will also change the frequency at which the antenna changes from a bi-directional pattern into a multi-lobed pattern.

To examine this question, I recreated the 27.2-m antenna model to perform frequency sweeps on both longer and

shorter versions. As a sample, I ran a 50-m version and a 15-m version. All of the models used 820-Ohm terminating resistors, #14 AWG copper wire, and a spacing of 0.2 meters.

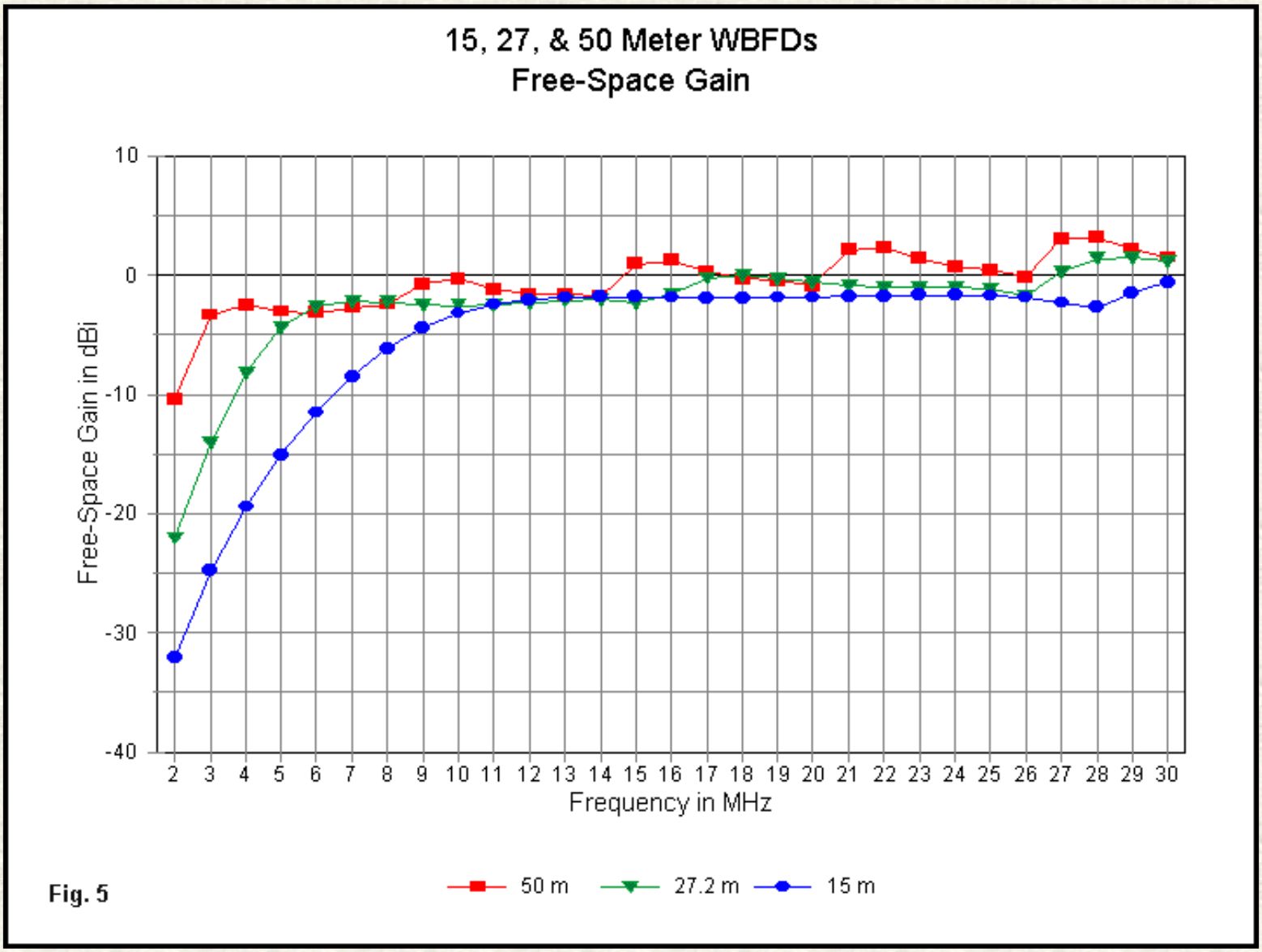


Fig. 5

Fig. 5 compares the gain of the three antennas from 2 through 30 MHz in 1 MHz steps. As suspected, the 50-m antenna reduces the knee frequency to about 3 MHz. In contrast, the 15-m version increases the knee frequency to about 10 MHz. In general, a home builder may interpolate values for the knee frequency for other lengths in the overall range.

The longest of the antenna models shows a mere -10 dBi gain at 2 MHz, a value easily made up by the receiver and only about 1.5 to 2 S-units below the average gain of the antenna. Hence, it is likely to be more satisfactory as a transmitting antenna in the lower HF region. In contrast, the 2 MHz performance of the 15-m version is more than 30 dB lower than the average antenna performance, making it more suitable for higher HF transmitting.

The variations in gain among the curves in the relatively flat region of performance are a function of lobe formation. Maximum gain tends to attach to the major lobes of patterns taken at just higher than integral multiple of a wavelength, relative to antenna length. Minimum gain levels tend to be associated with antenna lengths near the "x+.5" wl (where x is an integer) points. When an antenna is 1.5, 2.5, 3.5, etc. wavelengths long, its pattern consists of a combination of emerging and disappearing lobes, all of relatively equal strength. For example, a 1 wl wire has 2 strong lobes that are 180 degrees apart and a 2 wl wire has 4 strong lobes that are roughly 90 degrees apart. A 1.5 wl wire has 6 lobes, as

the 1 wl lobes diminish and the 2 wl lobes grow. Hence, coverage is wide, but at a reduction in maximum strength.

The number of peaks and valleys in the three gain curves is a function of length. The 50-m antenna passes through many more transitions from x wl to x.5 wl (where x is an integer) across the frequency span than do either of the shorter antennas. Hence, we should expect more highs and lows in the gain pattern.

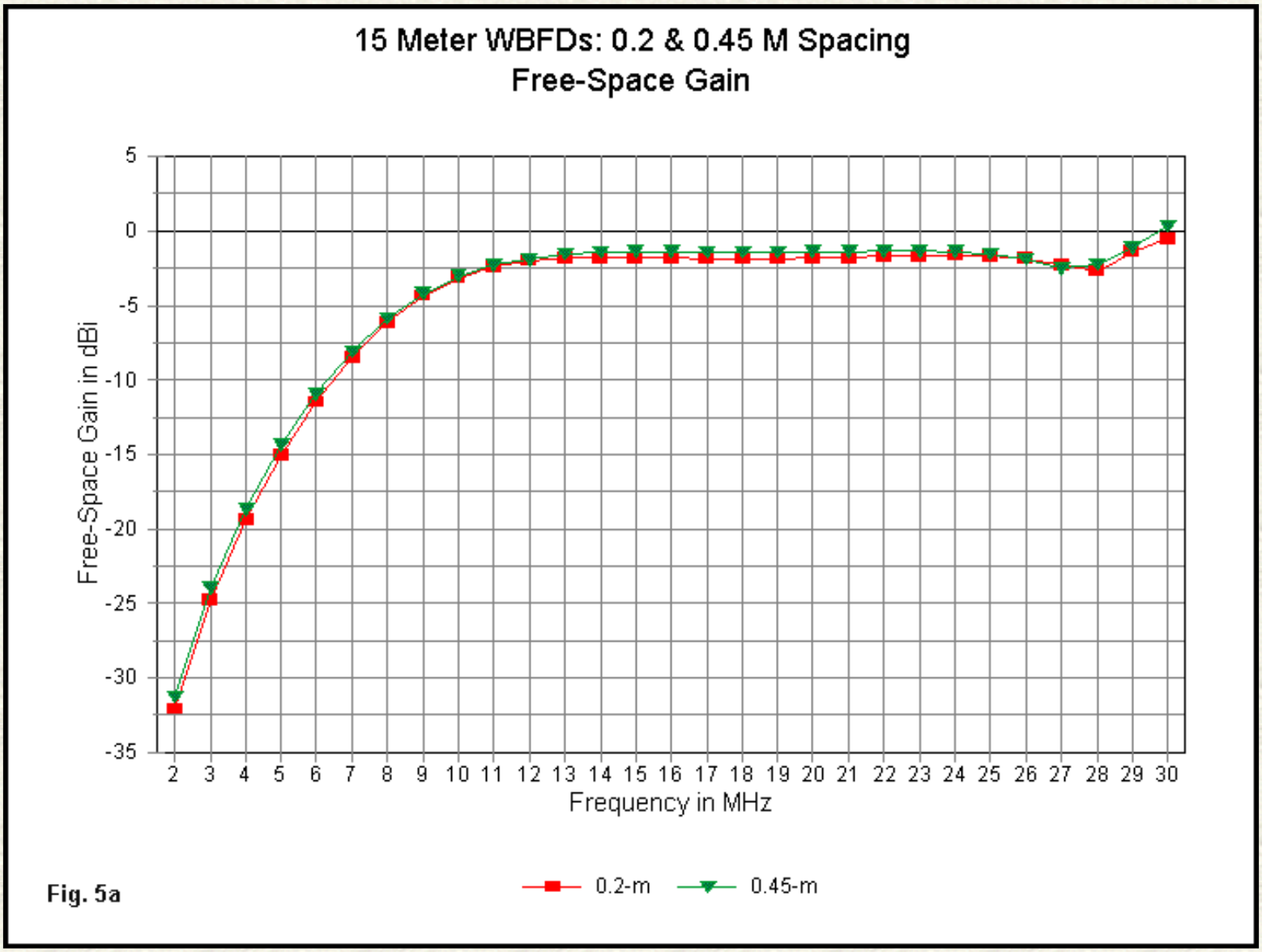


Fig. 5a

One question posed by various recommended wire spacings in past literature is whether wire spacing makes a difference to performance. **Fig. 5a** provides something of an answer as it compares the gain values for models of a 15-meter long version in 0.2-m and 0.45-m spacing. The gain values are insignificantly different, ranging from 0.2 to 0.4 dB.

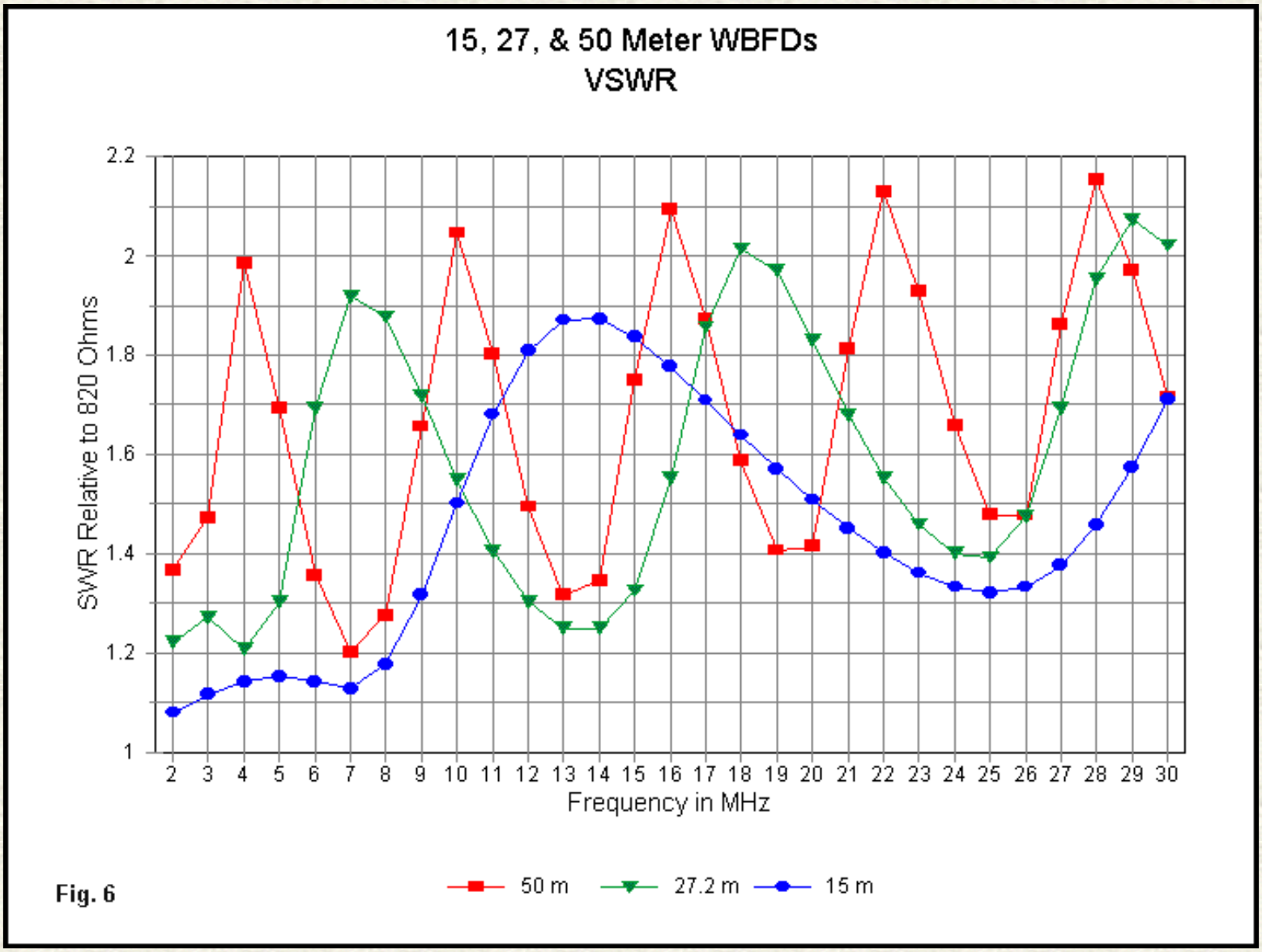


Fig. 6

We find that the curve of SWR relative to the value of the terminating resistor will also show similar transitions according to WBFD antenna length. **Fig. 6** shows the SWR pattern for the three antenna models. If we look at the most dramatic fluctuations--in the case of the 50-m antenna, we discover SWR peaks at x+2/3 wl points (where x is an integer). In contrast, we find SWR minimum values at the x+1/6 wl points (where x is an integer). The frequency span between points relative to the antenna length is 1/2 wl. The shorter antennas show the same pattern. However, the pattern is less evident because there are fewer maximum and minimum values to sample.

We may also note that the longer the WBFD, the higher the SWR excursions for a given value of terminating resistor. However, if we examine the lowest values of minimum SWR and exclude the region below the gain knee of the curve, the corresponding low points in the curve show the longest antenna also to exhibit the lowest minimum value of SWR. In other words, for a given wire size, spacing, and terminating resistor, longer WBFDs will exhibit a larger range within any given SWR cycle. As we approach the upper HF range, the values may exceed the desired 2:1 SWR limit.

The amount by which a long WBFD exceeds a 2:1 SWR is not great, but it is noticeable. For receiving applications, mild excursions beyond the 2:1 limit have virtually no affect on the received signal strength for any length of 50-Ohm coax. Some transmitters use automatic power reduction circuitry as the SWR approaches 2:1 (using an internal reverse voltage sensor), and some linear amplifiers begin reducing power at lower levels of SWR in order to protect expensive transmitting tubes.

There are two means of overcoming the potential problems of "high" SWR. Some manufacturers recommend the use

of very long coaxial cables. Since the losses in the line increase with frequency, the SWR observed at the station end of the line will be lower at higher HF frequencies than at lower HF frequencies for any given value of SWR at the antenna end of the cable. The result of using longer coaxial cable runs will then be an SWR curve at the transmitter output that never exceeds 2:1. Compared to the reduced gain already inherent in WBFD design, the added losses of a long cable run are not considered excessive when totalling the overall system gain.

Alternatively, modern amateur transceivers (and those in other services) are routinely (but not universally) equipped with automatic antenna tuner circuitry. Although limited in range compared to a wide-range external antenna tuner, these tuners are certainly adequate to handle the modest SWR values presented by even the longest WBFDs. Hence, the transmitter output circuitry prior to the tuner will show a very low SWR.

Construction

The decision to use a WBFD involves an evaluation of one's goals in operating or listening. Only with a set of specifications of this order can one decide whether the WBFD will meet the needs. The description of the antenna's advantages and limitations must be set against the operating specifications and along side other potential antennas that are candidates. Then selection becomes a matter of choosing the antenna that does most of the jobs well enough.

If you do decide to use a WBFD, you can purchase one of the commercially made types. B&W (USA), Giovannini (Italy), and others produce these antennas in a variety of lengths. Alternatively, you can build your own.

The antenna proper uses standard techniques of wide-spaced folded dipole construction. You will need twice the length of wire as you determine the antenna length to be. There is nothing critical about the exact length, although the general length will be a function of where you decide to place the frequency that forms the knee separating relative even performance at higher frequencies from diminishing gain at lower frequencies.

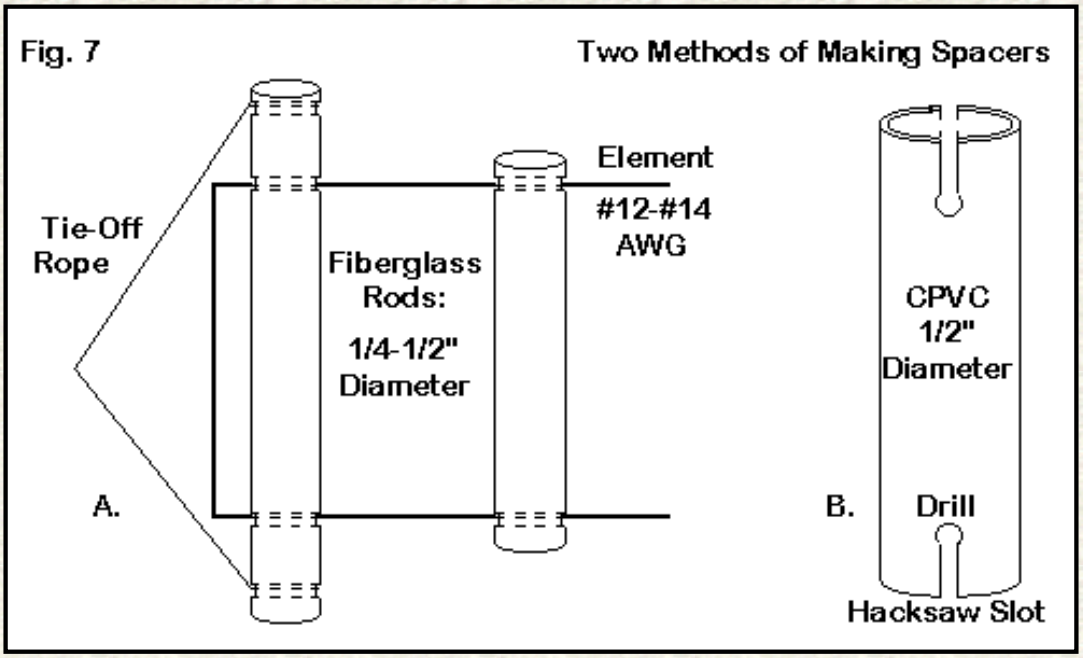


Fig. 7 shows just 2 of many ways to space the wires along their length. In the 1930s, we might have used wood dowels boiled in paraffin. Today, we have access to a variety of better materials. Part A of the sketches shows fiberglass rods, with holes drilled to pass the size wire we decide to use. #12 to #14 AWG copper wire (0.06-0.08" or 1.5-2.0 mm) is likely to be the most common choice. The end post can be longer to hold tie-off ropes for the assembly. Fiberglass rods can be purchased from mail order sources. However, local home improvement centers often carry adaptable materials. For example, I recently spotted some 1/2" diameter fiberglass rod under the guise of chimney flue brush extension handles.

Alternatively, I have also had good luck using 1/2" diameter CPVC, a thin-wall form of PVC tubing that replicates copper tubing sizes, shown in Part B of **Fig. 7**. A hacksaw cut in each end leads to a hole drilled to pass the chosen wire size. The wire press fits down the slot and into the hole. If the holes are not deburred, the wires stay put, although the spacers can be repositioned with fair ease.

These are simply two of many ways to make the required spacers. Narrow strips of polycarbonate, acrylic, or plexiglass would also work. Polycarbonate likely has the best UV resistance of this group. When adapting materials to a new use and environment, it is wise to check the structure every so often to ensure that it is wearing well under the influence of sunlight, precipitation, and temperature excursions. Of course, cut any spacers that you use to the desired length--about 8" (0.2 m) between wire holes for the models examined here. However, this spacing is not very critical.

Locating a non-inductive resistor of sufficient power dissipation is likely to be the chief problem for WBFD builders who intend to transmit with the antenna. Unless you can find a suitable resistor at one of the surplus outlets, purchasing an antenna may prove economical in the long run, if we add both cost and parts-searching time together. Any value in the 800-900 Ohm range--or even "thereabouts," if a bargain appears--will serve.

Manufacturers use different methods of packaging the resistor into the antenna assembly. Some prefer a total enclosure to weather-proof and bug-proof the resistor. However, one might have to derate the resistor's power handling capability under these circumstances. To maximize power dissipation, the resistor can be placed within a tube that is about twice the diameter and about 1.5 times the resistor's length. Air passing through the tube provides cooling, while the tube itself protects against immediate weather impacts. Since the antenna wire and resistor terminals will attach to strips of metal bonded to the tube, the resistor itself is relieved of strain. The down side of this technique is the need to clean out bugs and others debris on a regular basis. However, semi-annual inspection and antenna maintenance is always a good policy.

For receiving-only applications, the resistor problem is much simplified. A series-parallel combination of carbon resistors with a net value of about 820 Ohms is easy to arrange. 1 to 5 watt non-inductive resistors provide the sturdiest construction. The assembly should be mounted in a UV-resistant plastic housing with strong terminals for connecting the antenna wires.

The other challenging component is the 16:1 RF transformer. The builder has two general types of transformers to use: a transmission-line transformer or a standard wide-bandwidth transformer using a toroidal core. Transmission-line transformers are slightly more efficient for transmitting purposes, although they prefer purely resistive loads. Jerry Sevick, W2FMI, has written extensively on these units, with instructions on how to build them for many impedance transformation ratios. In a pinch, one might place two 4:1 baluns in series.

There are proponents of standard RF transformers using toroidal cores. Doug DeMaw, W1FB, has written on their use, including calculating the power-handling capability of various cores. For receiving-only applications, small cores can be used, and the basic requirements and calculations are described in recent editions of the *ARRL Handbook*, Chapter 6.

Whatever form of RF transformer you use, package it to withstand weather. A sealed UV-resistant plastic box with a correctly placed "weep" hole for moisture drainage is a good choice. Obviously, you will need connections for the antenna wires as well as a coax connector.

A 3-Wire Version of the WBFD

The WBFD has many variations, and from time to time--as I encouter and model them--I shall add a few notes on them. The first addition to the list is a 3-wire WHFD, outlined in **Fig. 8**, with the 2-wire version shown for contrast.

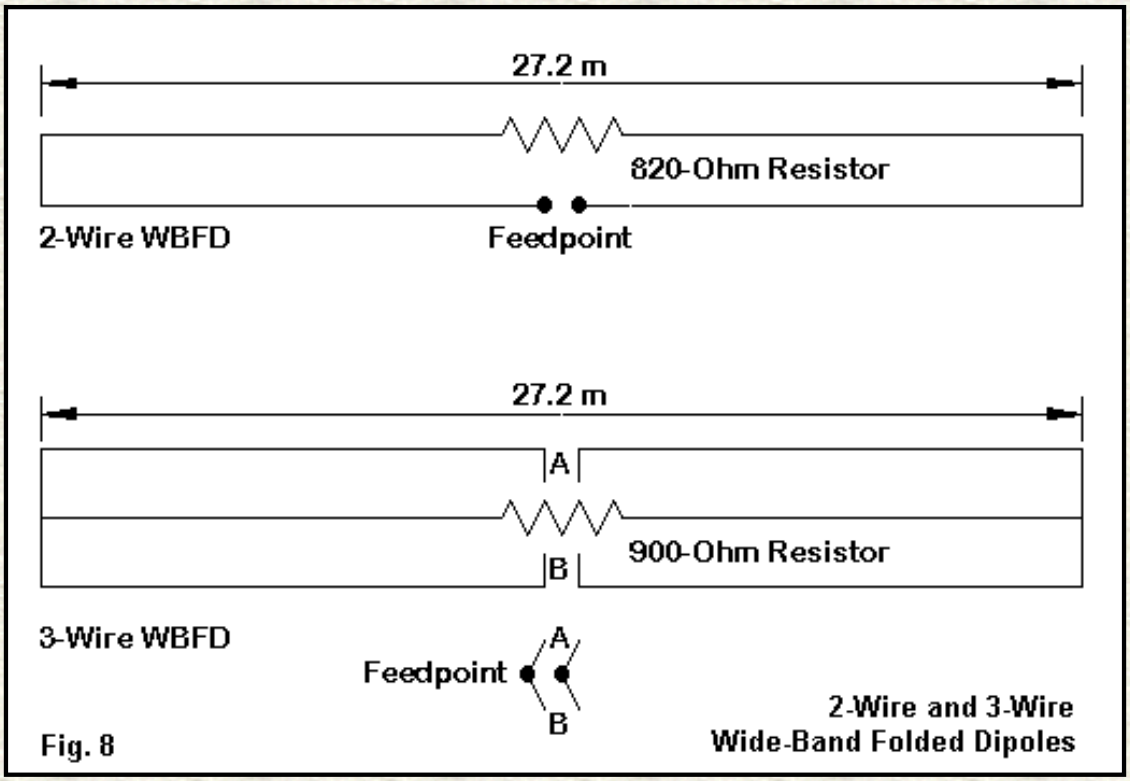


Fig. 8

For comparative purposes, I have made each antenna 27.2-m long, with a 0.2-m separation between wires. The 3-wire version places the terminating resistor in the "center" wire and parallel feeds the two "outer" wires. Although the arrangements is shown as a flat configuration, one can, as a variation, create a triangle of wires.

With two parallel-fed wires, the terminating resistor needs to be about 1.5 times the anticipated center feedpoint impedance. Hence, the model uses a 900-Ohm resistor, with 600-Ohms as the expected feedpoint impedance. Of course, the feedpoint impedance is a nominal value, since the actual resistive component of the impedance will fluctuate continuously above and below that value as we move across the operating span.

In general, the 3-wire version of the WBFD is capable of about 1-3 dB (depending upon frequency) gain advantage over the 2-wire version. The following brief table provides a glimpse at the fluctuations for the 27.2-m antennas. All gain values are for free-space.

Frequency MHz	2-Wire Gain dBi	3-Wire Gain dBi	3-Wire Advantage dB
5	-4.39	-1.62	2.77
10	-2.51	-1.11	1.40
15	-2.42	-0.39	2.02
20	-0.56	+0.74	1.30
25	-1.20	+0.53	1.73
30	+1.28	+2.58	1.30

Despite these fluctutations, the gain curves for the two versions of the WBFD are remarkable congruent, as shown in **Fig. 9**. The dual or parallel feed system of the 3-wire WBFD increases gain by feeding 2 wires, but it does not change the main characteristics of the antenna. Besides the congruence of gain curves, the patterns yielded by the 3-wire antenna differ from those of the 2-wire version only in peak gain, but not in strength. Since the gain increase is marginal, both antennas have patterns that replicate those of a simple doublet (with its widely varying impedance with changes in frequency), but remain smaller, that is, have much lower peak gain values.

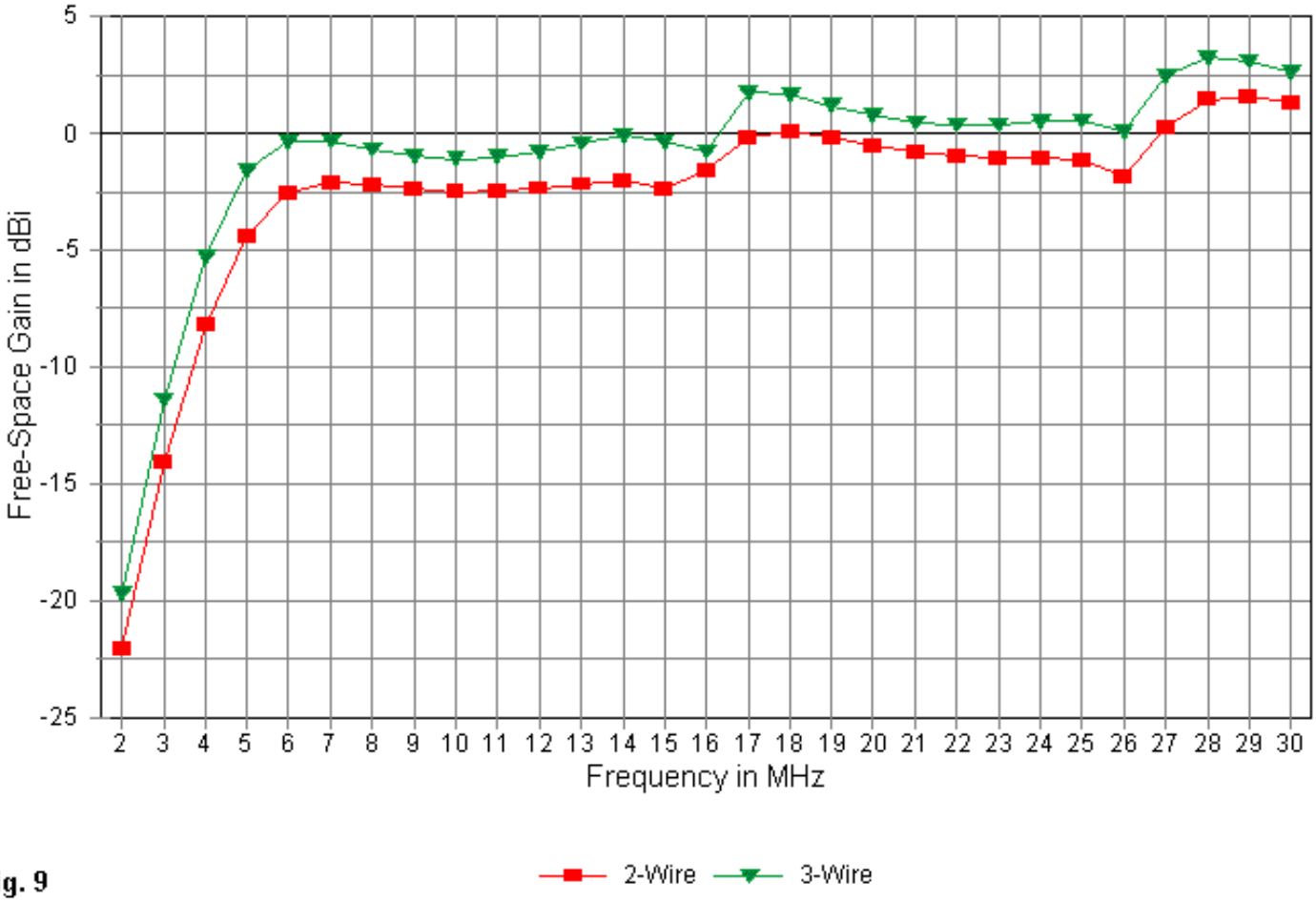
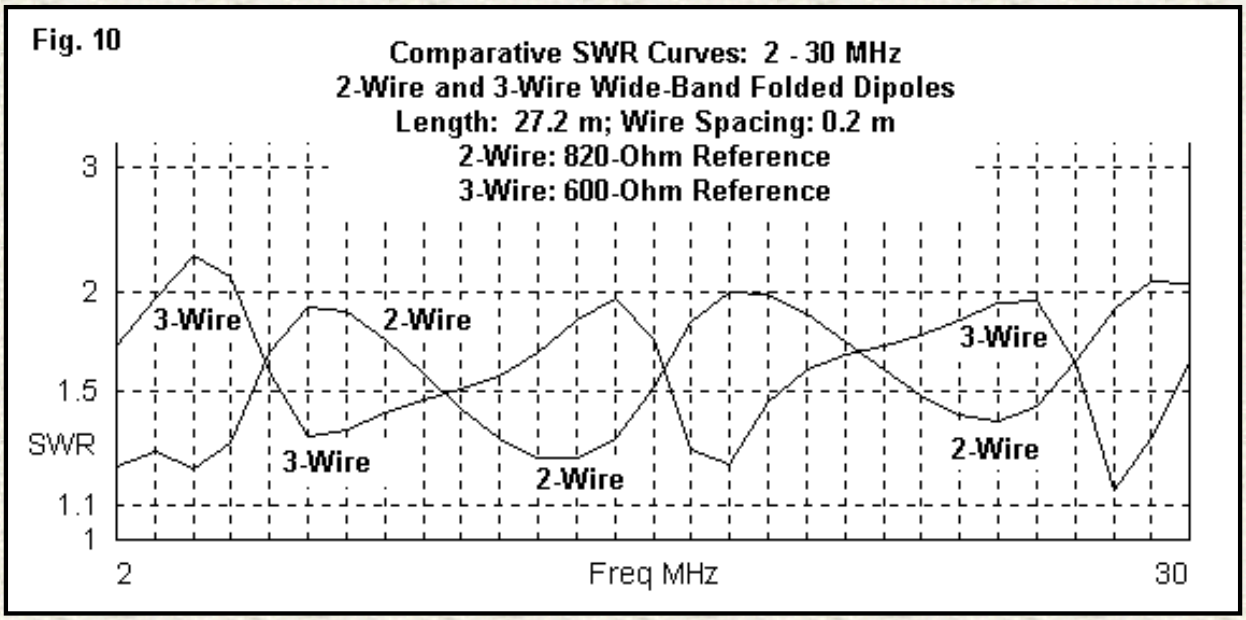


Fig. 9

The cost for the small increase in gain is a wider fluctuation in the impedance about a central impedance value. The reactance tends to be somewhat higher than for the 2-wire WBFD. **Fig. 10** shows the comparative SWR curves for the two antennas, with each one using its own reference value: 820 Ohms for the 2-wire version and 600 Ohms for the 3-wire version.



In addition to having higher peak values in the 2-30-MHz span, the 3-wire version curve slopes differ from the corresponding 2-wire slopes. The shallower parts of the curves are on opposite sides of the peaks.

The construction of a 3-wire WBFD can generally follow the same set of techniques used for the 2-wire antenna, with the separators enlarged to handle the broader plane. The additional wire will increase the weight of the antenna proper by nearly 50%, and center support of the terminating resistor and the feedpoint area is advisable. Short sections of 600-Ohm (or thereabouts) open-wire feedline can create the feedpoint junction. Like the 2-wire WBFD, the 3-wire version requires a balun system or a wide-band transformer to match a 50-Ohm coax feedline.

WBFDs can be used in inverted-Vee configurations. However, expect a decrease in the broadside gain and some vertically polarized radiation off the ends of the array--just as you would find in a simple doublet. The steeper the angle of the two side of the antenna, the greater will be the radiation off its ends. As the antenna length exceeds 1 wavelength, the patterns may increasingly differ from those of the antenna used horizontally.

Conclusion

A WBFD antenna is not for everyone. However, gaining some understanding of its operation, its advantages, and its limitations may be useful in the process of choosing an antenna--or even simply learning more about what various antenna types can do. The WBFD has its niche among amateur, governmental/military, and SWL antennas, but that niche is certainly not universal.

Receiving versions of the antenna can be home built for not much more than the cost of the wire, since the materials necessary for low-power terminating resistors and wide-band RF transformers are low. However, building a transmitting version of the antenna at home may be much more problematical, since parts may be hard to find or hard to fabricate. The alternative, of course, is one of the commercial versions, in an exchange of bucks for bother.

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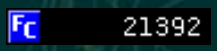


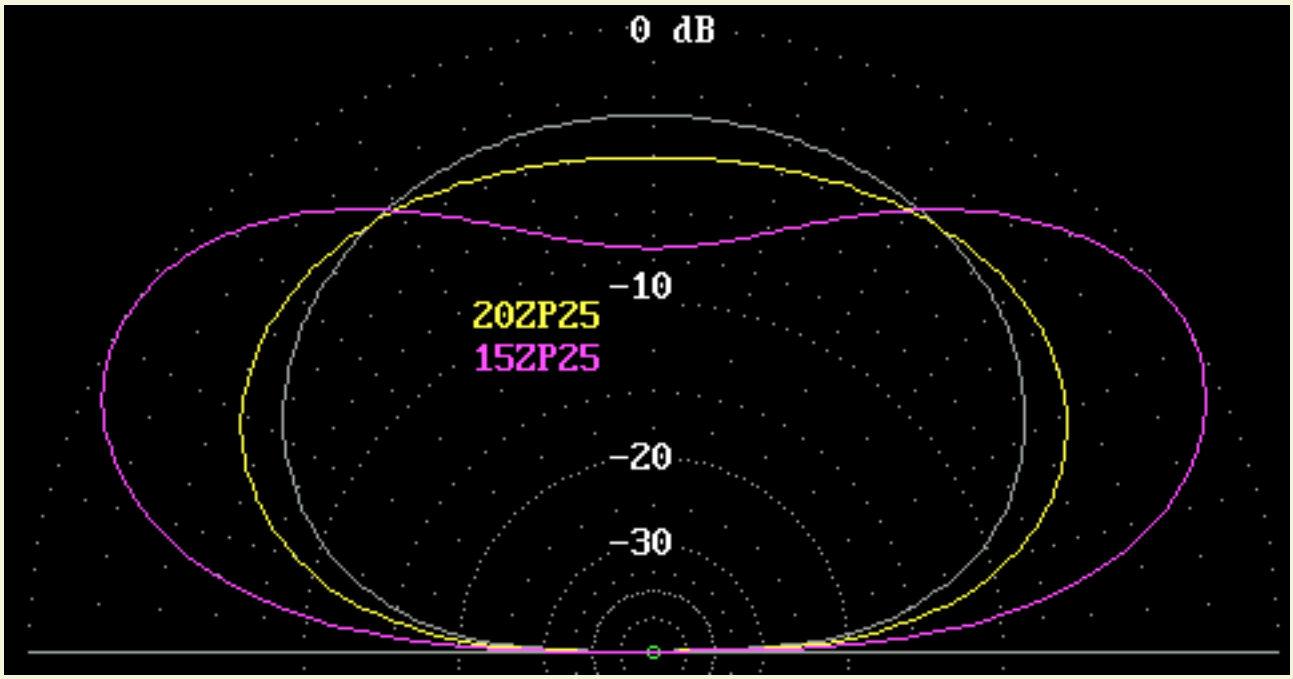
FastCounter by LinkExchange

The Extended Double Zepp Dipole

Cut for 15 Meters

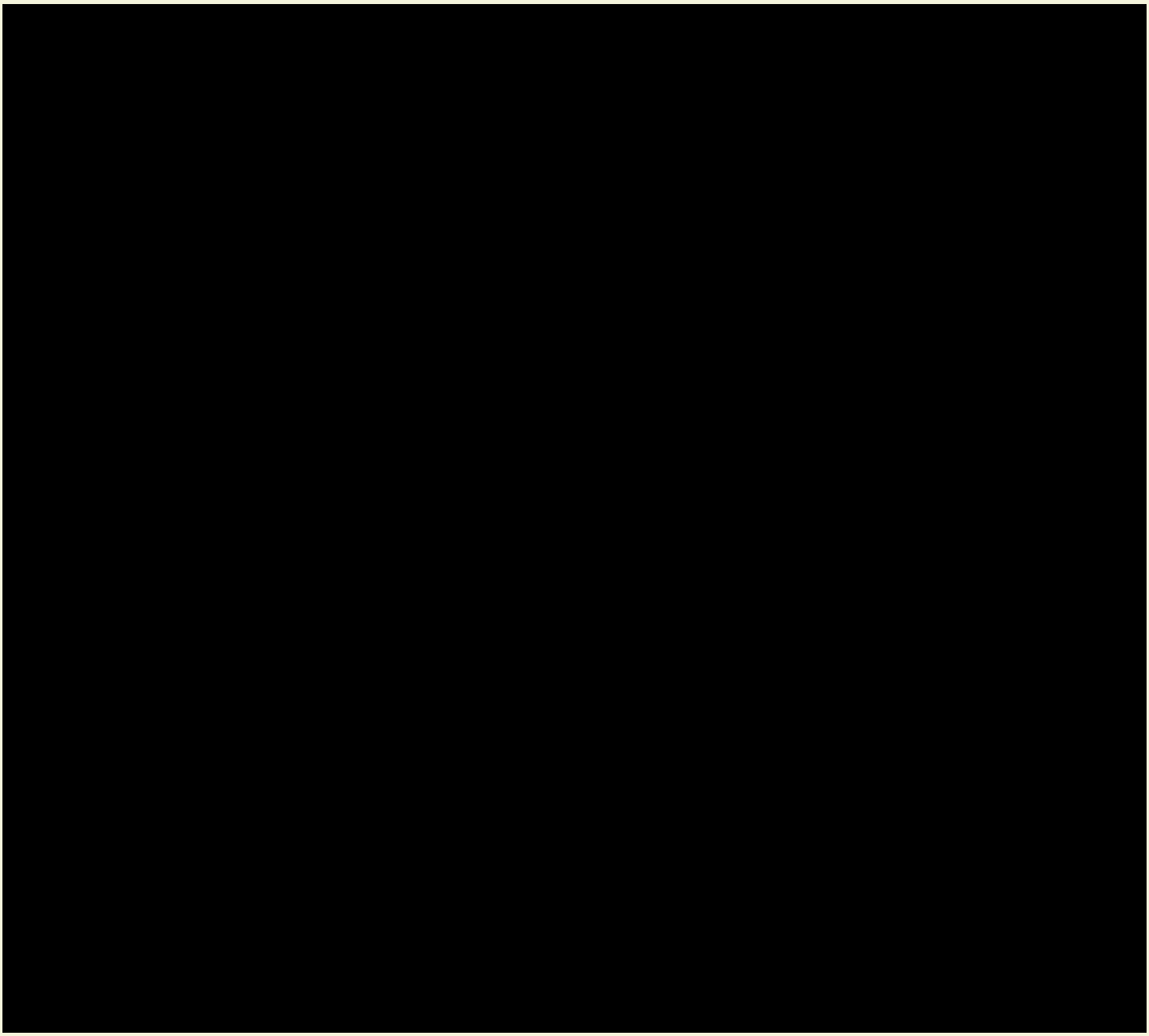
Mounted 24 feet high

An extended double zepp antenna for 15 meters is modeled mounted on two 24 foot poles spaced 60 feet apart. The antenna is cut for 15 meters and is 346 inches each side of center. It is feed in the center with 300 ohm twin lead. Typical feedpoint impedance is 119 ohms resistive and -692 ohms reactive. The EDZ is a fine antenna with several special properties.



Above we see a comparison of the broadside views of of the 15 meter EDZ loaded at 40 meters (grey), 20 meters (yellow) and 15 meters (pink). The 10 meter plot is not included here because it is split on the end as this approaches a long wire on 10 meters. Gain on 15 meters in the favored direction is 11 dbi with a launch angle of 26 degrees.

The EDZ is just about as long as a wire can get in terms of wavelength before the donut broadside pattern typical of a dipole quits sharpening and splits into two lobes. In fact you can see that already the split lobes are appearing at 15 meters. As frequency is increased further, these lobes grow rapidly and the broadside lobe shrinks rapidly. As wavelength is further shortened on a given wire length, it starts acting like a long wire with these split lobes getting closer and closer to the direction of the wire and multiple minor lobes appearing.



Viewed from above we see the 20 meter pattern (grey), the 15 meter pattern (pink) and the 10 meter pattern (yellow). Note how the 10 meter pattern has split and now has maximums that are 42 degrees off axis of the antenna broadside and an unfortunate null directly broadside to the antenna. This changes abruptly the favored angle of radiation of the antenna from dipole to long wire pattern. However, gain on 10 meters is excellent, being close to 12 dbi in the favored directions.

Another major consideration of the EDZ is mounting height. Since the dipole pattern is sharpened, it sends considerable energy towards the ground near the feedpoint. Thus the antenna changes pattern abruptly with height above RF ground. Optimum mounting height for such an antenna is close to .6 wavelength above RF ground. Below that there is some gain loss. Above that a vertical lobe starts to appear that saps energy from the desired low angle radiation.

This means an EDZ for 15 or 10 meters is a practical portable antenna. It can be loaded on the lower

bands, can be mounted only a 1/2 wave high which is just 21 feet at 15 meters and 16 feet at 10 meters, and it will perform adequately at lower bands while giving very worthwhile gain on the band it is specifically cut as an EDZ.

Of course the EDZ can be cut for lower frequency bands, but at 40 meters it now wants to be mounted about 70 feet high for best results. At 80 meters that climbs to 130 feet for optimum mounting height. Due to the fact the pattern will split just above the design frequency, the EDZ should be optimized for one of the higher bands you plan on using it on.

As long as you have a decent tuner designed to handle the reactive feedpoint which is highly capacitative and about 120 ohms resistive at the design frequency, you can effectively load an EDZ on many adjacent bands. The 15 meter one models as loading with a reasonable pattern on 40,20,17,15,12,10,6 meters. With the best results being broadside on 17,15 and 12 meters, though it works well on 20 meters.

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- MAGNET-LOOP ANTENNE, Berechnung, Konstruktion, Aufbauhinweise
- Duplexer (Diplexer) 2m / 70cm
- Duplexer (Diplexer) KW + 6m / 2m + 70cm
- Koax-Kabeldaten (Dämpfung, Kapazität, Abmessungen etc)
- Konversions-Tabelle zu Drähten (AWG / SWG / BWG / W&M / metrisch)
- Entstörung von Spannungsinverter 12 V auf 220 V
- Digimodes: Sauberes PACTOR - Adapterbox
- Strom-Balun, Mantelwellensperre

In ENGLISH

- WORLDS BEST HF MOBILE ANTENNA
- MAGNETIC LOOP ANTENNA, design, construction, hints
- Duplexer (Diplexer) 2m / 70cm
- Duplexer (Diplexer) HF + 6m / 2m + 70 cm
- Coax Cable Data (attenuation, capacity, dimensions etc.)
- Wire Gauge Conversion Table (AWG / SWG / BWG / metric)
- Clean PACTOR and digital modes - Adapterbox
- Current Balun

Zurück / Back HOME

The DX Notebook
Antenna Page

Antenna Links

- Antennas from QRZ Shareware
- ARRL Antenna Projects
- AA3PX Antenna Page
- Antenna Analyzers--G3YCC
- Antenna Height?--AA3RL
- Antenna Links--WA6BOB
- Antenna Magic--KQ6RH
- Antenna Theory 101
- Antenna Projects--AC6V
- VE3GK Web Site (Great Antenna Pics!)
- How to Become an Antenna Guru
- AntennaX Magazine
- Ralph Holland, VK1BRH
- FVR Spitfire Array for Top Band
- Four Square Vertical Phased Array -- YCCC
- FUNET Antenna Archives
- Rhombic Antenna Home Page
- K1TTT Technical Reference
- The Antenna Elmer
- SM0VPO/G4VVJ Homebrew Antennas
- Roofspace Antenna's from G3PTO
- Coaxial Cable Specifications from NA5N
- Antenna System Evaluator
- Cubical Quad Antenna Calculator
- Shortened Dipole de GM4JMU
- 160 metre "L" Antenna de N8SO
- How To Make Open Wire Feeders
- The Delta Loop
- How High Should my Dipole BE?
- AA3RL Transmission Line Calculator
- GAP Titan DX Evaluation
- EI7BA Multi-Band Quad
- The Crossed-Field-Antenna
- Antenna Visualization

Tower Service

- Rick's Tower Service

Manufacturers

- Aluma Tower Company
- Alpha Delta Antennas
- Antennas West
- Antennas & More
- Ant Panel Products
- Array Solutions
- Barker & Williamson
- Belden Wire & Cable
- Butternut & Bencher Antennas
- Bytemark (Amidon Products)
- Cable Experts
- CAL-AV Labs., Inc
- Comet Antennas
- Comtek Systems
- CushCraft Antennas

Antenna/Tower Mailing Lists, Forums & E-Zines

- TowerTalk Forum
- TowerTalk Reflector
- TowerTalk Reflector Archives
- The Condo Communicator -- a newsletter directed toward helping hams solve the perennial problem of getting on the air from inside apartments, townhouses and anywhere that the traditional big-tower setup is forbidden or impractical. Dormitory residents and anyone not living on 40 acres in the country will find the suggestions in these articles helpful.
- antennex Amateur Radio The Magazine All About Antennas PUBLISHED MONTHLY

Antenna Pictures

"Antenna farm" of 13-year old Mike DeChristopher, KB1FWN. The topmost antenna on the tower is a 15-element Cushcraft at 80 feet. The next antenna down is on an out-rigger to the left, which is also a 15-element Cushcraft, and even further down are two phased 2GHz antennas, one on the left, and one on the righ side of the outrigger. The beam on the bottom is a Bencher Skyhawk (When this photo was

- Cubex Cubical Quads
- Custom Metalworks / Rotating Tower Systems
- Davis RF Co
- Delta X-Ray Antennas
- Diamond Antennas
- DX Engineering Antennas and antenna parts.
- DX System Radio
- EM Scientific: "New" MININEC
- EUR-AM
- EZNEC/ELNEC Antenna Modeling Software
- Force 12 Antennas
- GAP Antennas
- GB Antennas
- Giovannini Antenna Systems
- Harbach Electronics
- Heights Tower Systems
- High Sierra Antennas
- Hygain
- IIX Equipment Ltd.
- International Technology Antenna

taken the reflector element was missing). The antenna closest to the photo is a Mosley TA33Jr at 36 feet.

Send your pictures!

- K3IWK's Chassis & Antenna Parts
- K0XG Rotating Towers
- KC2TX Antennas
- KN4LF Antenna Mfg and Sales
- KMA Antennas
- Larsen Electronics
- M2 Antennas
- LDG Electronics
- M2 Antennas
- Max Gain Systems
- MFJ catalog
- Mosley Antennas
- NEC-Win Antenna Modeling Software
- Nemal Electronics International
- PolyPhaser Corp.
- Radio Works
- Raibeam Antennas International
- Rohn Towers
- The R.F. Connection
- SGC World
- Steppir Antennas
- Titanex High-Tech HF

- Antennas
- Tennadyne Antennas
- Top Ten Devices, Inc
- Shoestring Antennas
- Sommer Antennas
- Texas Towers
- The Wireman
- TET-Emtron Antennas
- Trylon Towers
- US Tower
- Vectronics
- YagiStress

Please send comments, questions, suggestions or corrections to W5KNE

This document was last revised March 28, 2004

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www.dxer.org

DXpedition Antennas for Salt Water Locations

A study on 20m Antennas

By Kenny Silverman, K2KW

Which is better a 20m antenna for an island DXpedition: Yagi at 25', or a vertical or two on the beach? (Hint, You are on a web site dedicated to vertical antennas - this is a trick question!)

"Conventional Wisdom" would say that Yagi (typically a tribander or 2 element full-size Yagi) would be a great choice of antennas for a DXpedition on a beach. What makes the Yagi so great? Lets examine the reasons to use a Yagi…

The Top 10 Reasons to use a Yagi on a DXpedition, even when you are on the beach:

1. You have used Yagis at your home station
2. You understand how they work
3. Yagis have directivity
4. Everyone else uses Yagis on DXpeditions, so the *must* be great
5. and…

Hmm. I can't think of any other reasons! But everyone knows that *any* antenna by the ocean works great, or at least "conventional wisdom" tells you so.

O.K., lets examine the use of a Yagi for a serious DXpedition to a remote island. For this discussion, let's assume you are going on a DXpedition of a lifetime to Kingman Reef (KH5K) in the middle of the Pacific Ocean. Kingman is an interesting island which is about 25' wide, and 450' long. The entire island is made up of broken shells, rocks, and rubble. There is no sand, trees, or anything else. The wind typically blows at a steady 30 mph. You are days away from any kind of medical help.

Now imagine the difficulty of assembling a typical triband Yagi, where if you drop a nut, you won't likely find it again. Then your team has to armstrong the 35-40 pound Yagi, which is on top of a 25' mast, into the upright position. I'm assuming you aren't even going to consider a rotator, which would add another 25 lbs. to the top of the mast! Remember that the wind is blowing at least 30mph, and you are standing on loose rubble. To get the Yagi installed, you will probably need 3-4 people to walk the antenna up and hold it in position while securing the base in the shells, finding stable guy points (for 2 sets of guys, or 6 ropes total) in the rubble, and adding a tag line for rotating the Yagi. Any slip-up, and one could easily get cut on the rubble, or even possibly break a bone if you fell. Any relatively small problem like that in the tropics will likely become seriously infected in short order. Based on the wind and the hazardous conditions, you will probably need 4-6 people for a few hours to safely erect the antenna. This is looking like a lot of effort for just one antenna…Ahh, but waiting hams are worth the effort aren't they?

Maybe, but your health comes first. You may wonder - is there an easier and better way?

First, lets think a little more about propagation to Kingman Reef. Kingman is in the middle of no place, more precisely, in the middle of the Pacific Ocean. Based on the distances from the 3 main target areas, most take off angles are likely to be very low, usually under 10 degrees, and often below 2 degrees! Europe is the main target, and you will likely spend half your operating time on long path (typically very low take off angles). Most typical Yagi antennas have a beamwidth of around 60 degrees, which means that you will likely have to turn the Yagi for each of the main target areas:

Directions to target areas from KH5K:

JA: ~ 305 degrees

Europe: from ~355 degrees to 20 degrees (which is the main emphasis for the expedition)

USA: from ~43 degrees to 58 degrees

Continuing with our original "Yagis are the best" scenario, you remain convinced that the Yagi is still the best antenna. What if I told you that a simple 1/4 wave vertical on the ocean, or better yet a 2 element vertical dipole array, would be a far better performer! I can hear the pundits saying: "Verticals are for kids!" or "Real men use Yagis!" You believe there is no way that a Yagi could be inferior to a vertical!

So why do you think that verticals work so poorly? Because "conventional wisdom" says so? Unfortunately "Conventional Wisdom" on the performance of vertical antennas usually comes from comparing a Yagi at home to the trapped vertical in the back yard. At home the Yagi *will* be better, but not on the ocean! On the ocean, you will see that verticals are the clear choice for high-performance antennas, that just happen to be easier to install too!

First off, let me dispel one long-standing myth: Horizontal antennas over salt water do *not* get any enhancement from the salt water (from increased ground conductivity). Well, to be exact, almost all useful angles for HF propagation get little or no useful enhancement. The horizontal antenna (and so does the

vertical) receives a benefit from the ocean, because the ocean presents an undisturbed foreground for the incoming and out-going energy. Salt water also has less loss than typical ground for every reflection. To show you the impact of ground conductivity, the following two Figures were created: Figure 1 (below) shows that the take off angles in the 30-90 degree range (straight up) do get some limited enhancement from the salt water. Figure 2 (below) compares a 2-element vertical array over land and salt water. The only antenna that gets *significant* signal enhancement by being next to, or over, salt water is a vertical. In fact, verticals get about a 6 dB increase of gain when placed over salt water, and the radiation in the pseudo-Brewster angle is filled out (which is radiation under ~12 degrees). So the salt water is enhancing signals right were most DX signals are arriving - in the pseudo-Brewster angle!!

Figure 1: 2-element Yagi at 25' over land and salt water. Peak gain is 10.40 dBi (over salt water) at 38 degrees take off angle in this design.

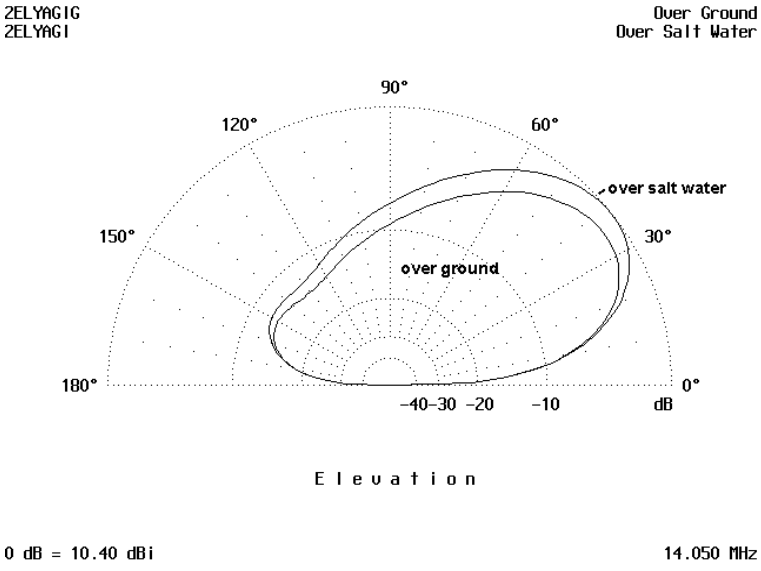
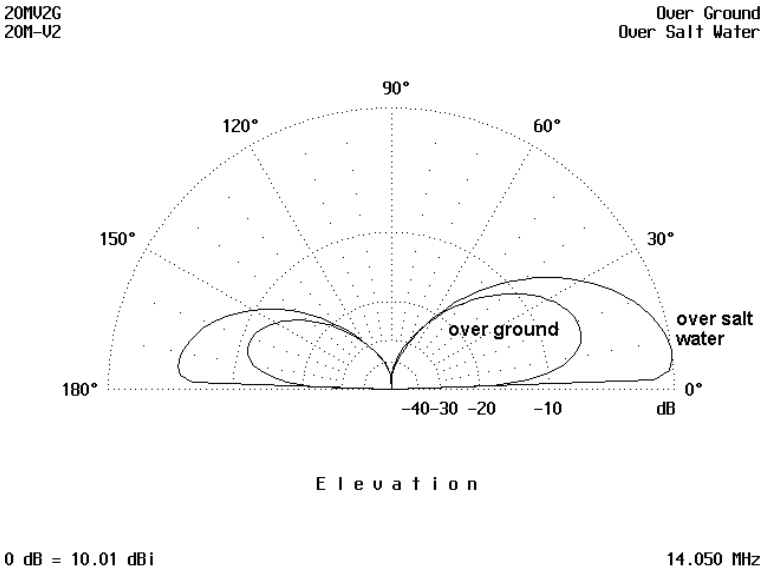


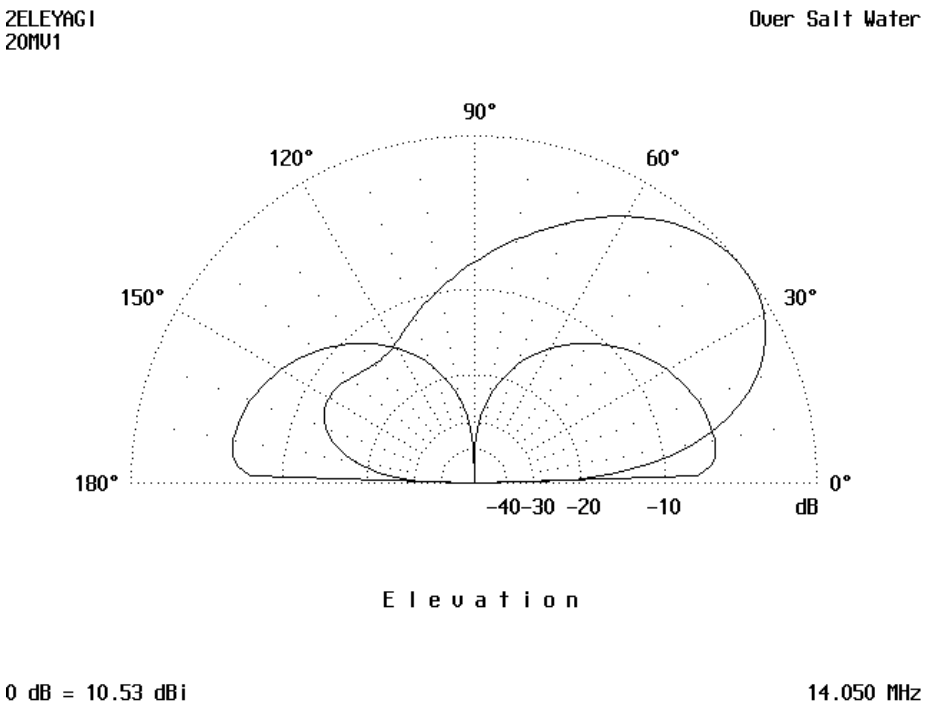
Figure 2: 2-element vertical dipole array (parasitic) over average land and salt water. Peak gain is 10.01 dBi (over salt water) at 8 degrees take off angle in this design.



OK. Let's get back to Kingman.

I suggested that a simple vertical would be better for most useful take off angles… let me show you why. Figure 3 (below) compares a 2 ele Yagi at 25' height over salt water compared to a 1/4 wave vertical over salt water. The antenna with lots of gain at 40-degree take off angle is the 2-element Yagi at 25'. The other antenna is the 1/4 wave vertical. Notice, at take off angles below 10 degrees, the vertical is the hands down winner! You may point out that the Yagi has more "gain". This is true, but the extra dB or two gain is at take off angles that don't matter! A simple vertical by the ocean can and will usually outperform a Yagi on most typical DXpedition paths, because the energy from the vertical has fewer hops to the target! It's all a matter of the angles of the arriving signals.

Figure 3: Yagi at 25' compared to a 1/4 wave vertical over salt water

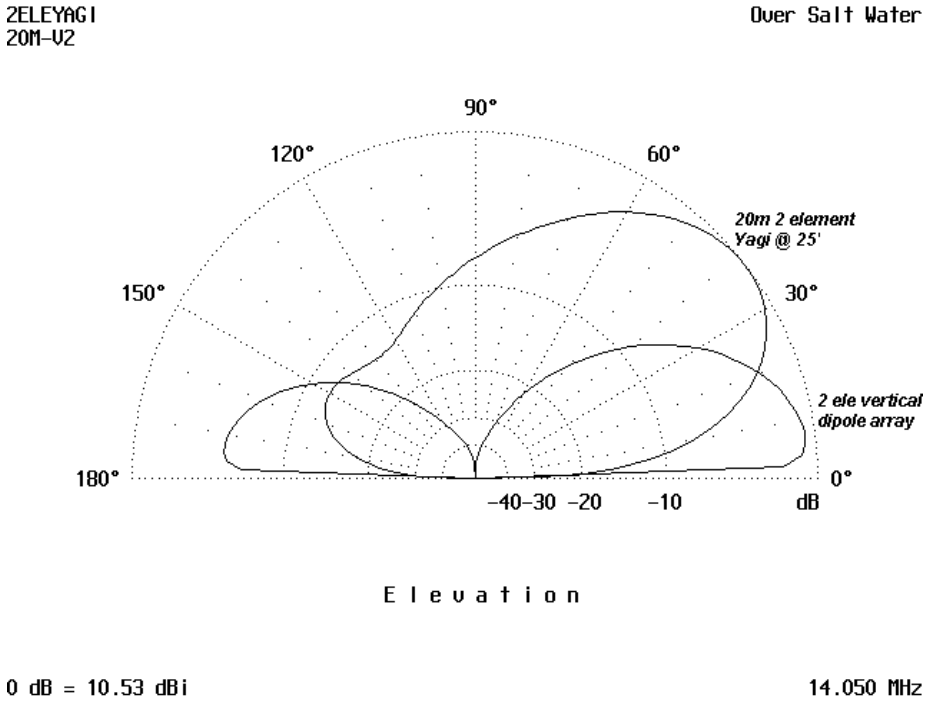


As you saw in Figure 3, the 1/4 wave vertical is a good performer. Even better for our Kingman Reef expedition is that a single vertical is really light weight (maybe 3-4 lbs for the 16' vertical), and can be assembled and installed in just a few minutes. Compared to the Yagi, the vertical only needs some rudimentary guying, and you don't have to worry about turning the antenna! And since you are going there to operate, you can be on the air in 30 minutes if you wanted! Try that with the Yagi.

I bet that some of you aren't convinced yet that a single 1/4 wave vertical is a good enough choice. OK, lets up the "vertical anti" a bit, and rather than a 1/4 wave vertical, lets examine a 2-element parasitic vertical array using 1/2 wavelength vertical dipoles. To better understand what this array looks like, picture a 2 element Yagi standing vertically, but minus the mast and boom to support it (much less overall weight). (see picture to right)

There are a few advantages of using vertical dipoles (vs. 1/4 wavelength elements): they eliminate the need for radials, and by raising the feedpoint, you achieve an additional 1.5-2.0 dB of gain at low take off angles. A parasitic vertical array (vs. a phased array) is an easy way to increase gain and directivity, yet is still very easy to install and tune up. The parasitic array needs less parts than phased array, which is an important aspect on DXpeditions. Figure 4 (below) compares the 2-element Yagi at 25' to the 2-element vertical array using vertical dipoles. You can see that the vertical array now has nearly the same peak gain as the Yagi, but the gain is focused right where most of the arriving signals are coming from: 0-10 degrees! And the verticals are now the clear winners from 0-20 degree take off angles - who could ask for more? Since many of the signals are arriving at around 1 degree take off angle, often the verticals can be 20 dB stronger than that Yagi! Talk about a band-opening antenna!

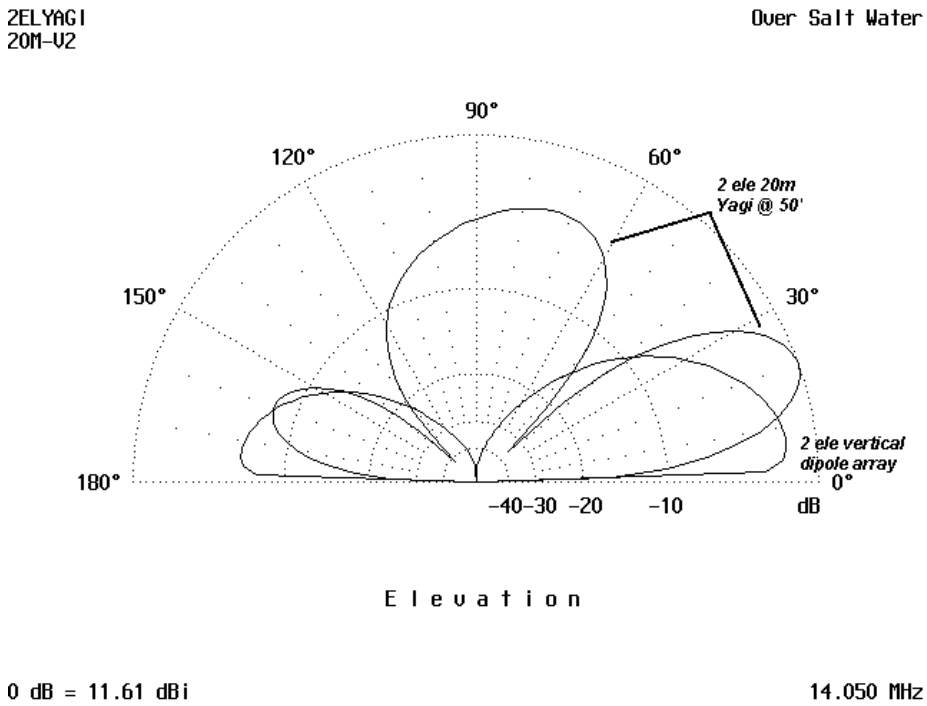
Figure 4: Yagi @ 25' compared to a 2-element vertical array using vertical dipoles



But I'm sure that the Yagi-lovers are saying: "I bet if you were able to get the Yagi up to a good height like 50', that it would surely beat the vertical array. Think again dude! First off, installing a Yagi at 50' on a DXpedition is difficult, if not impossible. Secondly, it still doesn't compare to the vertical array! Figure 5 (below) compares the 2-element 20m Yagi at 50' to the vertical array. Notice that the Yagi now has a large amount of energy being radiated straight up. Sure the main lobe is lowered, but the 2-element vertical will still be better on most DX signals.

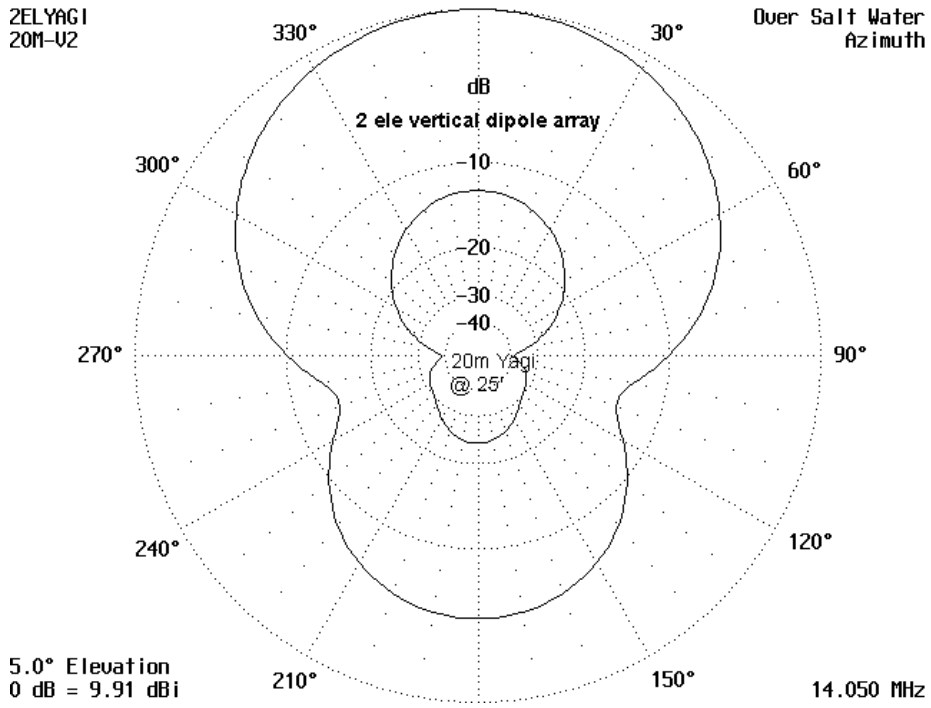
Remember, that a typical tribander weighs at least 35-40 lbs., and add another 20 pounds or so for the mast, for a total of at least 55-60 lbs. A two element full-size vertical dipole array for 20m weighs no more than 18 lbs. if properly built. A vertical array is much more efficient if you consider a key metric for DXpeditions: dB per Pound of Antenna.

Figure 5: Yagi @ 50' compared to 2-element vertical array using vertical dipoles



And you still need more convincing? OK, Figure 6 (below) is the energy at 5-degree take off angle. The antenna with all the signal (the big one) is the 2-element vertical array which has 9.9 dBi at 5 degree take off angles. The antenna that is down 12-20 dB in all directions is the good ole Yagi at 25'… Another key point is that a 2 element vertical array has a very wide pattern, which is on the order of 120 degrees for the -3dB points. Thus the 2-element vertical array can cover all 3 main target areas without turning the antenna!. In addition, if there are callers from other directions (assumes there is a water path in those directions), the vertical has more useful gain than the Yagi does on azimuths that are off the side or back of the antenna.

Figure 6: Energy at 5 degree take off angle of a Yagi @ 25' compared to a 2-element vertical array using vertical dipoles



Some of the pundits might still say this was a "Made up" example…Sorry to disappoint you, but the Kingman Reef example was real, and was based on the planning discussions for the upcoming Kingman Reef DXpedition. For many island DXpeditions, vertical arrays should be given more consideration on the high bands than they currently are receiving.

The computer models have been verified by empirical testing we did from salt flats while shooting over salt water. The measured data follows the model. Verticals were selected for the "Team Vertical" locations in the Caribbean and comparisons to full-size Yagis were made, confirming the models, as well. The comparisons were done over many continuous hours of switching back and forth, with differences of up to 9 S-units (on an 'MP meter), averaging 2-4 S-units, in favor of the verticals. The Team Vertical Yagis were occasionally used during the contests, but only to cover directions where the verticals has nulls. This might be one of those situations where if you haven't tried it, you don't realize what is happening. How good is a larger vertical array using vertical dipoles? Tom, N6BT, has commented that they are truly on par with the large commercial curtain arrays he used in Saipan. Salt water is the key to verticals. If you can get close to salt water, or literally sit right in it, such as at Kingman Reef, verticals are the answer to high performance antennas - besides being significantly easier to transport and set-up.

73, Kenny K2KW

Special thanks to Tom Schiller, N6BT, for offering input to this article.

eh-antenna · EH Antennas

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Files
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Links
Database
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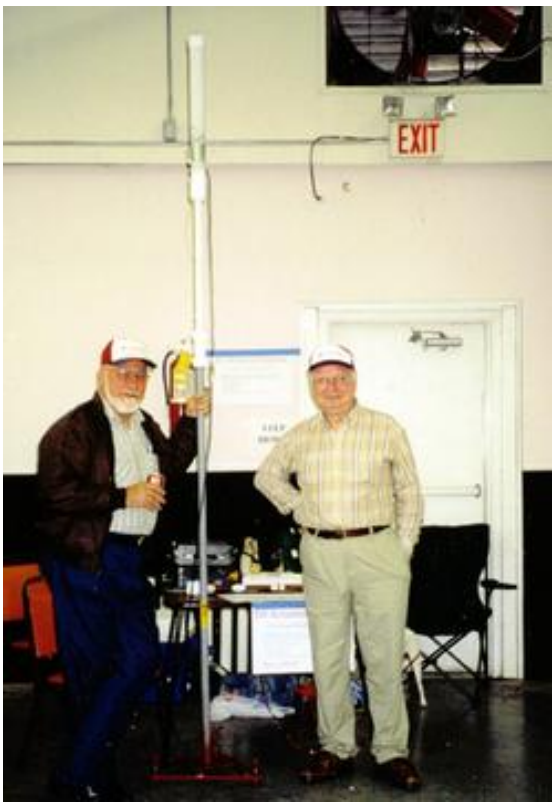
Calendar
Promote

(Already receiving group email?)

Description

Category: Amateur and Ham Radio



This group has been created to promote discussion relating to the EH Antenna. There should be many with experience with the EH Antenna, so don't be shy about asking questions on construction and efficiency.

There is a file and picture section, feel free to use them.

If you wish to Un-subscribe, use yahoo groups to un-subscribe just like you subscribe, the instructions are there, use them.

DO NOT SEND A MESSAGE TO THE GROUP ASKING SOMEONE TO DO IT FOR YOU

Picture of Ted (on the left) with me selling the Backpacker Kit at the Dalton, Georgia Hamfest.

What's New

New within the last seven days:

Members: **3** Messages: **18**

Message History

	Jan	Feb	Mar	Apr	May	Jun	Jul	Aug	Sep	Oct	Nov	Dec
2004	272	66	171	89	108	125	38	45	18			
2003	142	267	493	518	433	411	249	141	130	145	105	429
2002											188	134

Group Info

Members: **659**

Founded: **Nov 11, 2002**

Language: **English**

Group Settings

- Listed in directory
- Open membership
- Unmoderated
- All members may post
- Archives for members only
- Email attachments are permitted

Group Email Addresses

For more information: http://eh-antenna.com

Post message:	eh-antenna@yahoogroups.com
Subscribe:	eh-antenna-subscribe@yahoogroups.com
Unsubscribe:	eh-antenna-unsubscribe@yahoogroups.com
List owner:	eh-antenna-owner@yahoogroups.com

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Ham Radio - mySimon: Ham radio on mySimon. This great comparison shopping site offers you excellent product and pricing information across the Web plus a directory of links.
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E-H Antenna

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Background

Many antennas function because of common mode currents, rather than working in spite of them. Two popular examples are CFA and EH antennas. Another recent example, appearing in Antennex's compact antenna articles, is a thick stub "vertical" with no counterpoise.

All of these antennas become significantly poorer radiators if common-mode currents on feedlines are eliminated. Why? Because the feedline is the actual radiator, ***NOT*** *the tiny thing they call the antenna.*

Misunderstanding or misapplying Maxwell's equations and the principles behind radiation, in combination with missing some very key points of conventional circuit theory, causes problems. Some of us have unwittingly attributed increased feedpoint resistance and/or seemingly disproportionate amounts of radiation from very small structures to new methods of radiating EM waves. Reviewing these antennas and the theoretical or technical mistakes surrounding them will help us understand how antennas and transmission lines work. With that knowledge, we can build better antenna systems. The fastest and best way to learn is often to look in detail at mistakes!

What They Claim

Articles and user reports of CFA, EH, and thick stub verticals (without groundplanes) appearing in Antennex and other internet publications have one common thread, the operational descriptions almost always include strong indicators of problems with feedline common-mode current.

Authors commonly warn users to NEVER choke feedlines with baluns and to "be sure the feedline is straight and in the clear"! Authors lay blame for RF burns from feedlines or shack equipment on the antenna's "high radiation efficiency", claiming these small magical antennas radiate so efficiently they naturally excite the feedline and equipment more than full-size antennas.

For example, Ted Hart (of EH Antenna fame) claims the following at:

http://eh-antenna.com/documents/RF_on_the_Coax.pdf

"RF on the Coax

Due to the large radiation at the EH Antenna, there will be some RF coupling to the coax. Whether this is a problem is dependent on the radio you use. Some are subject to RF coupling into the audio system, which causes severe distortion while transmitting. On some field day setups with 100 watt transmitters we have had so much RF on the radio you can get an RF burn. Below we have suggested ways to eliminate the RF coupling problem."

In the above statement, Ted actually acknowledges current on the feedline shield and RF-voltage-on-chassis problems. The problem must be severe when a low-power 100-watt radio causes a burn. Like any good salesman, he turns a design shortfall into a feature! According to the author, unwanted RF on the feedline doesn't come from a feedpoint or antenna design problem like it does on other antennas, in this case it appears because the antenna works so well!

Here's what actually causes RF to appear on a coax shield and radio chassis. RF can only appear on the radio chassis through two methods:

1.) The antenna, from poor feedline or feedpoint design, can couple to the radio chassis through external wiring or cables attached to the radio.

2.) The radio chassis itself, being large in terms of the wavelength, can actually become an antenna and receive energy from actual desired "over the air" signals.

(Many of us have these problems. Click on this link to see one reason why.)

In this case, we can probably rule out reason two above. It is unlikely the chassis is a large portion of a wavelength long on 20-meters and that the antenna field is suddenly so strong it is "lighting up the house" with RF. After all, if power is radiating effectively it will all be going out to distant stations and NOT cooking you or the radio gear in your house! That only leaves reason one, poor feedpoint and antenna design, as the cause of common-mode feedline or wiring currents that excite the coax shield and eventually the radio chassis, as described in method one.

Actually someone has measured this, and posted it to a web page!

http://www.home.earthlink.net/~calvinf15/_technical/

When the time-varying current from the transmitter flows in any conductor, we will have charges accelerating in the cable. The outside of a shield is no exception. A feedline's shield will radiate proportionally by the ampere-feet of the cable. just as any other conductor will. Of course the antenna element will radiate also, but there is something else to consider.

A very small current flowing unopposed over a large linear distance will radiate quite a strong signal, because radiation resistance of a long antenna is generally very high compared to very short antennas. You can find this explained in the Radiation&Fields and Radiation Resistance articles on this site, and in engineering textbooks such as those written by Jasik, Kraus, and Jordan-Balmain. From all of this, we know the shield radiates.

The inventor of the E-H antenna goes on to say:

"If you use RF beads, since the coax shield is not a magnetic shield, the beads affect both the inner and outer conductors. Therefore, most of the transmitter power will be converted to heat. Not good."

Not a very knowledgeable statement at all, at least from the standpoint of how shields work!

Time-varying fields can not pass through a shield that is more than several skin depths thick. The inner part of the shield and outer part are isolated by the skin effect, and nothing passes through. The ARRL Handbook, Maxwell's book "Reflections", Reference Data for Radio Engineers, and dozens of other amateur radio and engineering texts describe this effect correctly. If we bring a time-varying electric field to zero in a system, the time-varying magnetic field is also by definition zero. The shield DOES isolate the center of the cable from time-varying magnetic fields on the outside of the cable!

If that is true, why then does a shielded cable passed through a current transformer used in a directional coupler appear to pass RF magnetic fields? Why does the RF magnetic field seem to "pass through" the shield of a shielded receiving loop antenna? The answer is quite simple. There is a gap or intentional break in the shield.

Current on the inside of the shield "spills over" the edge of the shield where the shield is broken, and causes a current on the outside of the shield. There is also a voltage across the gap at each end of the shield. We have both our time-varying voltage and current, via a circuitous path to the shield ends! Our Amateur texts explain that effect, as do all of the engineering texts dealing with shields on transmission lines. If the gap is closed and the shield's ends are shorted together, making voltage across the shield gap zero, the magnetic field no longer "seems to" penetrate the shield!

If a shield did not behave this way, we'd be in serious trouble in the radio world. Without the shield stopping both magnetic and electric time-varying fields, we could not shield our radios. We also could not shield our microwave ovens, with non-magnetic materials!

If anyone thinks a ferrite core affects the impedance inside a coaxial cable that does NOT have significant common mode current, they only need to slip some beads over a cable on a working normal "cold-for-RF" feedline. You will find absolutely no difference in system performance when

beads are added, proving Ted does not "have it right" in the text above that appeared on the E-H antenna web site.

The author and inventor of the E-H goes on to say:

"Use of a small choke made of several turns of the coax is good. We find that a wire connected to the ground side of the coax at the antenna and connected to either a ground rod or a wire laying on the ground will eliminate RF problems - in most cases. For some radios we also need to add a ground wire to the radio."

Of course adding a ground wire might help! The ground wire becomes the path for common-mode currents, or at least a portion of them. The additional wire to ground becomes part of a long-wire antenna (made by the cable shield) that actually does most of the radiating!

"A preferred method is to run the coax to ground then back to the radio. Near ground, connect the shield of the coax to a ground rod or radial. Another method is to connect a wire from the radio to ground. If the radio is very far from ground you will need to add a series resonant circuit in the ground wire to effectively cancel the inductive reactance."

It is understandable why this is a preferred method! The outside of the coax shield can remain the primary antenna, saving us the bother of installing the additional ground wire that becomes the antenna in the previous suggestion!

"It may take one or more of the above to solve your problem. Remember that if you have a good ground on the antenna, you have also minimized problems with lightning."

In the above text, we can see every solution carefully avoids installing an effective choke balun on the feedline. A properly designed and effective choke balun has no effect on a coaxial feeder or system SWR, unless the feedline is radiating!

Let's look at another E-H antenna experimenter and former proponent of the E-H antennas test of a 160-meter E-H antenna. You can read the text directly at:

http://www.qsl.net/iz7ath/web/02_brew/18_eh/english/pag11_eng.htm

Steve writes:

"With my short (and easy) tests I deduce the EH lose about 10 dB versus my short 10 mt vertical with capacitive hat in the top;
10 db is a big amount of power, but remember we are using 2 small cylinder of 37,5 cm on 160 mt band; my vertical performed better than a 40 mt long sloper, so, may be, loss versus full size dipole is not too much."

Steve claims the E-H antenna looses about 10dB compared to a 1/4wl sloper, which has an unpredictable efficiency. A typical sloper 1/4wl sloper is likely only around 50% efficient in the best situations, and more likely much worse! The actual range can be from 10 to nearly 100% efficiency. Steve's data repudiates Ted's claim and the CFA inventors claim that "crossed-field" antennas provide high efficiency.

"There seems to be something good;
May be optimizing something we can have more gain (or, better, we can lose less dB);
This antenna can be useful on the low band, expecially on 137 KHz (the EH cylinder should be 6 mt high with a 2mt diameter).
On the web I've seen an other similar antenna (ISOTRON), wich seems caming (as EH Antenna) from CFA antennas;
On the italian Radio Rivista (1995-1996) there were same informations about an "Antenna Toroidale" wich in same way remember the EH: will be interesting build and test one."

Steve seems to be saying something we all agree with. A short antenna will radiate, but not nearly as well as a full-size antenna.

"By the way, I think that my EH is not for Dxing; it will be useful for that radioamateur who have no possibility to install "long" antennas for low band, but want just to have local QSO or few contest-contact."

I agree! 10-20dB loss from a full size antenna would not make a good DX antenna! Now here come the current-on-the-feedline problems:

"Other point: tuning;
I think it's not as simple as others say; I've used an MFJ259B which tells me all about the antenna, but the tuning was critical. On the other side construction is easy;
I want suggest to you , when tuning, to connect the MFJ259B to the shack ground (if there's one). If not, EH will probably resonate higher (20-30 KHz)."

If grounding and ungrounding an antenna analyzer or any other piece of equipment connected to a coaxial cable causes resonant frequency or SWR of an antenna to change, the system has severe feedline radiation problems. See my article on testing baluns.

"Coaxial cable influence: inserting more coax cable, the resonance seems to vary a little; I suppose that's normal;

I don't know for certain if the line radiates or not; I've added two iron-powder coil forms back my TX and EH performance was unchanged. Then later I've added 2 big ferrite beads with a lot of turns on the roof (at the feed point of the EH) and all has changed. My signal was 3 S point weaker than before and S.W.R was very high.
I don't know if that loss is due to the changed resonance in EH (as I said before, outside the bandwidth performance goes down quickly)."

We see again that any attempt at reducing feedline radiation results in an antenna that does not radiate very well. With the feedline choked, Steve's EH antenna dropped 3 S-units in addition to the original 10dB from his sloper, or maybe 25dB or more!

There surely is a hidden message in all of the above contradictions!

How the E-H Antenna *Really* Works

There are many examples where designers intentionally use common-mode currents. Examples are found in textbooks, such as the "Antenna Engineering Handbook" by Jasik on and around page 22-6.

The antennas at the right, copied from Jasik's textbook, outline the derivation of a skirt collinear antenna from a simple feedline with the open end terminated into a conductor. (It could be a ground

Two or even three skirts may be added to the mast as shown in Fig. 22-5c, but reduced excitation of the lower sections diminishes the effectiveness of each additic skirt.[6] Thus, while the multiple-skirt coaxial antenna resembles a collinear arra in-phase half-wave elements, its gain is not as great. Typical gain values are ta lated in Table 22-2, but since no test data are available for this type of antenna, values given are merely engineering estimates.

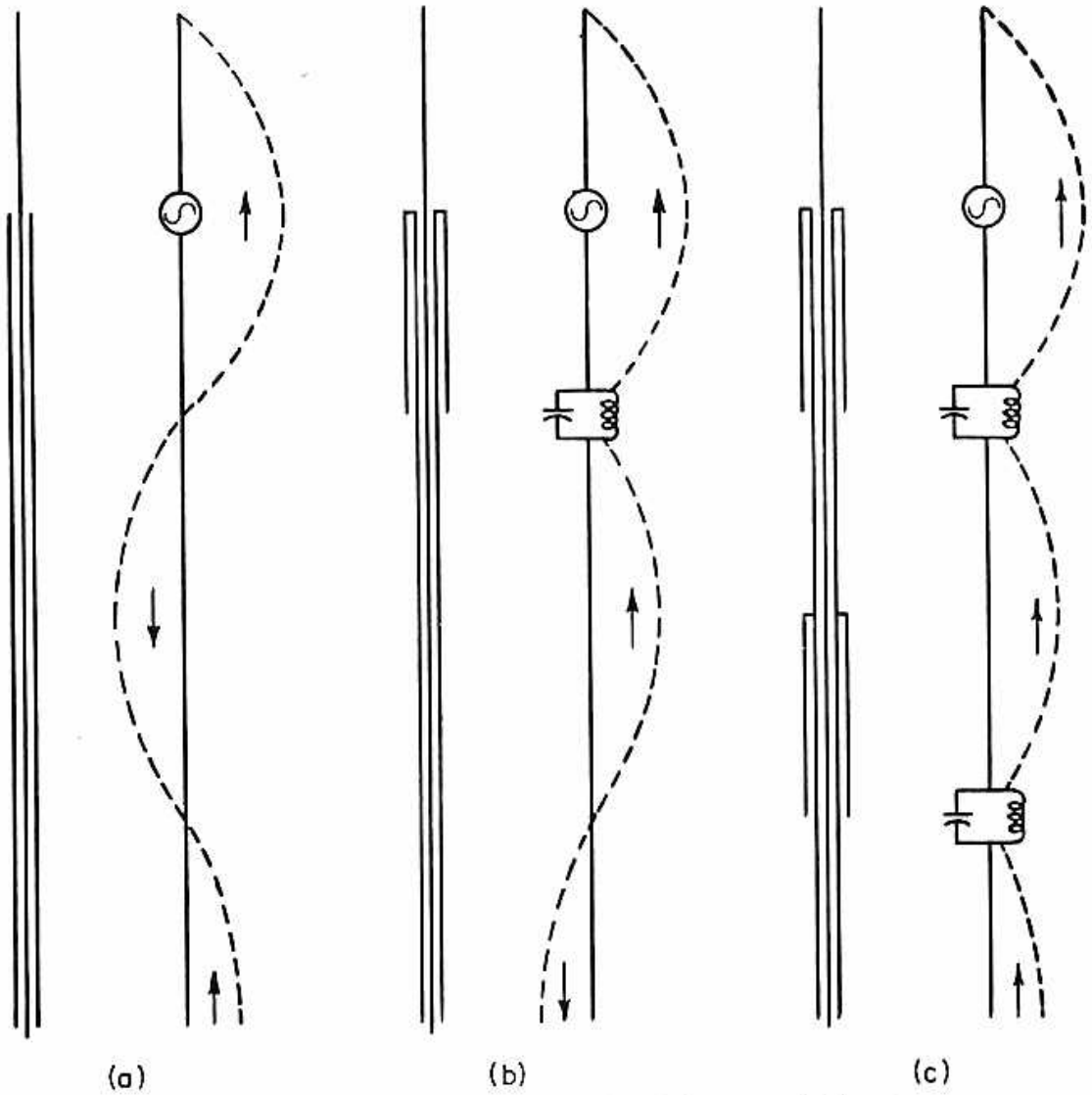


FIG. 22-5. Evolution of multiple-skirt coaxial antenna.

The relatively large diameter-to-length ratio of a practical skirt produces an effect which requires that the exterior length of the skirt be reduced by a factor of 0.8 to 0.9 and the interior length by from 0.95 to 0.98.[6] This is accomplishe cutting the skirt to the proper exterior length and inserting a dielectric slug ir skirt to increase its electrical length.

rod or an antenna, like a Beverage or large loop, the antenna does not have to be an "open circuit".

Looking at (a), we find by hanging any conductor from the end of a coaxial cable the shield is excited (on the outside) with common-mode current. The electrical equivalent of the OUTSIDE of the shield is just as if a generator located at the end of the shield was driving the outside of the shield as longwire antenna. This goes along with Kirchoff's Laws, that tell us the sum of currents entering a point must equal the sum of currents leaving that point. For any current to flow up into the antenna, and equal current must flow back down over the outside of the shield.

With one ampere flowing up the center conductor into the "stinger" at the coax's end-point, the same level of current flows back over the outside of the shield. (The shield's inside and outside are isolated by the skin-depth of the current at the operating frequency, an can be treated as two independent conductors that are connected over the open edge of the shield.) We *MUST* have this current simply because this is how coaxial cables work, the current on the inside of the shield is *ALWAYS* equal and opposite to current in the center conductor. There has to be some place for that shield current to flow, so it makes the bend over the end of the cable and flows back down the outside.

This is also why, when we use a cable's shield as a ground lead the center conductor and inside of the shield do nothing to reduce resistance. Any current that flows down the center conductor is cancelled by current flowing on the inner wall of the shield, the result is no current at all flows down the center conductor as long as the shield is several skin depths thick.

Many antennas intentionally and unintentionally use this principle, two examples are shown in (b) and (c).

A recent Antennex Article on a "magical" ultra-compact antennas claims an identical system, using a loaded fat cylinder, has an extremely high radiation resistance and excellent performance because some magical field-trickery increases the radiation resistance of a thick cylinder at the antenna end. Certainly radiation resistance is somewhat high...but not for the reason something magical or special is happening!

The small coil-loaded cylinder is actually only a fraction of the antenna length, and being so short has a very small radiation resistance. The point missed is the shield of the cable is very long, and is in series with that short section. Since the shield is long, it also has a reasonably high impedance both from radiation resistance and loss resistances. Shortening the length of the end-stub results in an insignificant reduction in radiation resistance, because the overall length of the radiating system is very long! We have a simple off-center fed dipole, with one very long leg and one very short leg!

The main radiator is the outside of the feedline shield, not the tiny thing being called an antenna!

Unless we make the coaxial shield an infinite length or pass it through what amounts to an infinite groundplane with zero resistance, current continues on down the cable shield. Looking at (c), we find even multiple sleeves appearing as parallel tuned high-impedance circuits do not fully decouple the

shield. Many collinear antennas work on this principle, yet E-H antennas and others attribute it to some form of electro-voodoo.

Jerry, K5OE: Home Brew Amateur Antennas

Home

LEO Sat. Antennas:
- ***Eggbeater II***
- Eggbeater
- ***TX Potato Masher II***
- Texas Potato Masher

Portable Antennas:
- 70 cm Handi-Tenna
- 2m Mobile Vertical

AO-10 Antennas:
- 70 cm Quagi
- AO-10 Beam

AO-40 Antennas:
- **Patch Feeds**
- **Dual Patch Feed**
- **Helix Experiments**
- **MMDS Dishes**
- **S-Band Toys**
- **2' Dish Mods**
- **Drake 2880 Mods**

Miscellaneous:
- **Satellite Links**
- Russia 2000
- Home Brew Hardware

Picture Album

The *Texas Potato Masher II* Antennas at 10 m elevation

Welcome to my Home Page (best viewed at 1024x768). I have a few "home brew" antennas here as well as some links below to other sites that may be of interest to the amateur builder/experimenter. I hope you find something of value here. Please send me any comments or suggestions to at (insert my callsigne here)@amsat.org.

If you would like some information on amateur radio operation, please visit the American Radio Relay League (ARRL) for an introduction to this wonderful international fraternity. If you would like information on amateur satellite operation, please visit the Radio Amateur Satellite Corporation of North America (AMSAT-NA) for the latest in news and software. If you are new to amateur satellites, here is an introduction to the "EasySats" (slightly dated, but still useful). If you are wondering what kind of antenna(s) you might want or need, please read this introduction to amateur satellite antennas and this LEO satellite antenna cost/benefit analysis.

I live in Houston, Texas; about 90 km from the Gulf of Mexico (New! UPDATE: I am temporarily in Darwin, NT, Australia, operating as VK8OE (APRS). Weather here is hot and humid, just like Houston. My QTH in Houston was in a neighborhood with "deed restrictions," outlawing antennas of any kind. My response to this situation was to install "low visability" antennae that are hidden from the street and do not extend above my roof line. Please see my various antennas for examples.

HF Antenna Pages of Interest:

W4RNL Antenna Design & Modelling

TowerTalk Reflector (Contesting.com)

AC6V Home Brew Antenna Page

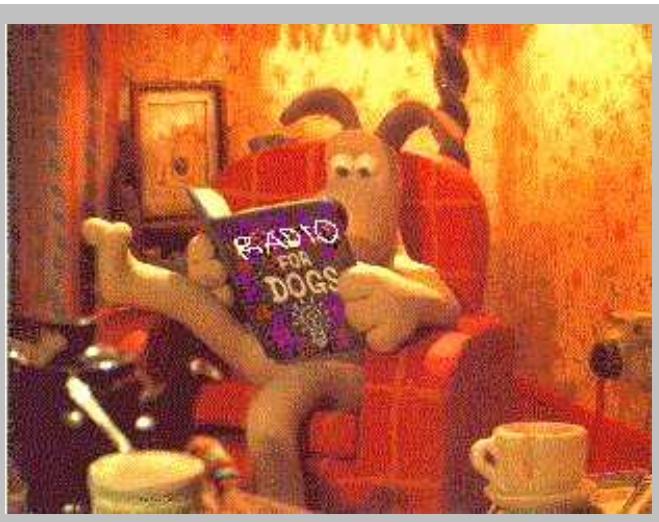
Satellite Antenna Pages of Interest:

Quadrifilar Helix Antenna

Lindenblad Omni Antenna

W1GHZ Microwave Antenna Book

KK5DO's AMSAT Net Antenna Script



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73,
Jerry, K5OE/VK8OE
EL29ds/PH57km

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The Gallery of the EME Arrays.

A Collection of EME Reflector Antennas and YAGI Arrays.

28-ft Dish at K2UYH.

The Majority of these pictures were provided to K2UYH for the 432 MHz And Above EME Newsletter.

Many of these antennae have been Designed, Constructed and Optimized by Radio Amateurs.
The Antenna represents a Major Module in a Moonbounce and/or Weaksignal System. It is an essential part in the receive as well in the transmit link.
A 1 dB increase in antenna gain delivers 1 dB increase in ERP and, depending on the frontend, it could deliver a more than 1 dB increase in system S/N ratio.
A significant improvement in an EME link Budget!

Antennae from the 432 MHz and Above Newsletter

- **K1RQG's dish under construction.**
- **DF3RU at 432 and 1296 MHz.**
- **W2DRZ and K2TXB 3 cm, a new dish is under construction. EME**

- K3HZO at 432
- Details of polarization and rotation at K3HZO
- OE9FKI at 5730. MHz
- OZ9AAR 5 meter dish with W2IMU feed.
- W4NJP 36 ft dish.
- W5UN, the Mighty Big Array.
- K5GW's 432 array.
- EA3EM, 3MD, 3UM, 3AQJ, 3AYX, 3BTZ, 3DXU, 3EHQ, 9AI and I5WBE.
- F6KSX and F1EHN and their 23 and 3 cm EME Antenna.
- F5PAU Dish.
- Dishes for 432, 1296 and 10 GHZ at F6CGJ. Worked WA7CJO and SM4DHN on 3 cm.
- 10 GHz at JA0IXX
- 2 x 16 x 22 elem K1FO's for JH0YSI. 16 Hor.+ 16 Vert.
- JH3ERQ, JH3EAO, JR4BRS and JA4BLC
- DJ9YW, Heinrich 4.5 meter, f/D=0.47, W2IMU Circular Feed. 360 W HPA.
- 6 Yagi's at JH4JLV.
- N2IQU's 48 ft, f/D=.41, inner 36 ft. covered with fine mesh.
- Sunrise in the Low Lands, PA0JMV.
- My good friend, PA0AVS and his upgraded Low Noise Parabolic Horn for 432 and 1296 MHz.
- 3.7 meter at HB9HBU.
- G3WDG de VK3ALU, 3.7 meter. Box behind dish with TWTA, Relays and Xverter.
- G3LTF, 20 ft Dish with 70 cm Feed installed.
- JS3SIM with 4 25 element K1FO Yagis.
- JR9NWC 8 Yagi Array high up in the blue Sky.
- KB2AH and his antenna park.
- Chuck, W8MQW with 8 K1FO yagis at 432 MHz and a 10 ft dish on 1296 MHz from the country side.
- W8MQV with 20 Watts on 10 GHz
- The Black Forrest uW Moon Bouncers, DJ7FJ(r), DL2GSG and their 4.5 m ANDREW Dish. Rig under dish.
- OK1DFC with 4 * M2 Yagi's.
- JA2TY, 4 * 33 element K1FO and FHX06 LNA.
- Takuhei, JA2TY running 2 * 3CX800A7.
- 5M Dish for 23 cm at DF9QX.
- PA2CHR Array with 4 * 35 el Yagi's.
- JA5OVU's new 16 * 14 el Rear Mounted Yagi Array.
- VE6TA, with 4 * 22 element K1FO and a 3.7 m, expanded to 5 m, on 1296 , .3 F/D.

- IK0EQJ, active on UHF and EME with bay of 4 Yagi's.
- The N7ART 1296 feed article and the modification of this feed by VE1ALQ
- VE1ALQ's modified N7ART dual band Feed
- The feed,...the feed at VE1ALQ
- Is this the Focal Point or Not? Darrell, VE1ALQ
- Antenna Maintenance at ON5RR
- G4ERC, Rotary Yagi Array
- HB9SV and his Dish for 23 and 13 cm
- JA7BMB, 6m .5 f/D Dish, with feedmounted water cooled 7289 PA
- JA4BMB's feed and frontend
- W2VVV, 8 * 18 el M^2 Yagis on Ring Rotator and M^2 El Rotator
- DL6UW, 6 meter on 1296 MHz, the shack is in the 10 ft high army container
- DL6UW Dish plus mount
- RW3BP, 10 GHz EME, from an apartment in Moscow
- Close Up of Feed, LNA and HPA of 10 GHz antenna at RW3BP
- 2 * 28 elements on a 9 wavelength boom at IK5QLO
- SV1BTR's 432 MHz Cross Yagi Array
- OH2DG, constructing a new Dish.
- QRP on 1296 MHz, TM8EME de PA3DZL.
- W6HD, 6 m Dish and 80 Watts at feed.
- DH9FAG, 3 m TVRO dish with 200 Watts and 0,35 NF.
- Dom, HB9BBD with his large dish for 1296 (details later)
- WE2Y and his YAGI Array on 432 MHz.
- 9H1ES, 2 m Dish for 1296 MHz.
- JH4JLV and his antennas.
- 6 * 36 el Long YAGI Array at DL3FI.
- The LU1296 Moonbounce Team, L-R LU6DW, LU4DHD and LU8EDR.
- Rotable dish feed by G3LTF.
- HB9SV, 16 Yagi 432 Array.
- KB4CNI, 8 * FO33 approx 28 dBi gain.
- HA5SHF with his 1296 antenna with feed .
- TVRO Dish at HA5SHF for 1296 MHz EME.
- N6BQ antenna party with WA6KBL and members of the 50UP group of N. Cal
- F1AHN, 1.8m Dish with 800 W at feed.
- OZ6OL in the shack.
- Dish at OZ6OL.
- Closeup of 432/1296 MHz feed at OZ6OL.
- W1ZX with a new and bigger dish.

- K6JEY, 4 * 25 el K1FO, ,4 dB, 500 Watts
- N2UO with his 1296 MHz dish
- **Room for More Antenna Pictures.**

Antennae uploaded by FTP:

- Shukon, 7M2PDT on 432 MHz with 8 x 25 elm. K1FO, TS790 and 8938 with 1.3 KW out.
- Scott, AC3A with 4 * M2 9E1 in a suburban , no towers, development.
- CT1DMK, Luis, 5.6m Dish for 1296 and 10 GHz.
- Guentner, DL4MEA on EME.
- 4.5-m Russian Radar Dish Modified into 6.4-m Dish for 23 and 13 cm at OH2AXH.
- Ian, G3SEK and his 432 MHz 12 * 15el DJ9BV array with Rotatable Polarization.
- G4CCH, 1296 MHz dish antenna in yard
- G4CCH, watercooled 1296 HPA.
- G4CCH with a 5m homebrew Dish Project.
- G4DZU's 3m Dish, with VE4MA Scalar Feed and 500 W on 1296 MHz.
- G4DZU's in front of the Dish.
- Toshi, JR4BRS, 5.6m Dish for 23 and 13 cm.
- KB3PD, Yagi Array for 144 and 432 MHz.
- KD1DU, Del with 4 yagis for 144 MHz EME and much more on the tower
- KE7MK with 14 ft TVRO Dish, KB2AH feed and final with approx 100 W.
- Scott, KD4LT and his reflectors for 70 and 23 cm.
- Doug, K6JEY, with 4 * 25 el K1FO, .22 dB NF, and 1200 W from DM03wt
- Monica, KQ6PY tightening the hub bolts of a 5m Dish (photo by WA6KBL)
- Jeff, K7XQ with a bay of 4, just above the roof.
- Anders, LA8LF, with home built Dishes.
- ON5OF and his antennas
- Luc, ON6JY running a smaller Station.
- N6BQ, Ron on 1296 from the Bay Area
- NU7Z, 4 meter Dish for 1296 MHz through 10 GHz.
- Frank, W2UHI with 5.5m Dish at 23 cm.
- Don, W4RDI, 5.5 m, .5 F/D.
- SM2CEW, 8 m, 0.6 F/D, 432 and 1296 MHz, mesh-size good for future 2304 MHz.
- F3VS, 6 x 4 20 el DL9BV, **submitted by Lean Marc Lair from REF magazine.**
- SV1BTR, 4 x 11 wavelength, Cross H/V pol BVO on a non conductive boom.
- Ron, N6BQ on 1296 with a TVRO dish.
- DF4PV, Guenter with extended 6m TVRO Dish and 600 W on 1296.

- **S57RA, bay of 4 YAGI's from the backyard for 70 cm. An anti aicraft gunmount as support.**

- **Parallel Wire Feeding of A Yagi System.**

Your antenna picture is welcome here as well.
Upload by Anonymous FTP to: nitehawk.com/pub/upload/rasmit/

Another fine collection of EME antennas by Skukou, 7M2PDT

More EME Newsletter Content from Tom, K2AH

Technical Articles and Information from Newsletter

- **K2AH shows the construction of 6 tube 23 cm Amplifiers.**
- **1296 MHz Power Amplifier by K2AH.**
- **PreAmps and PA's by Tom, K2AH.**
- **From the Newsletter: Tom, K2AH on Tuning 1296 MHz Feedhorns.**
- **Just for fun. [Java Enabled Browser required]**
- **Simple preamps for 144 and 432 MHz by IK5QLO.**
- **HB9BBD with moonsignal recordings and a mighty fine Dish.**

More Antenna information found at links:

- **Elevation Solutions for Yagis By GM4JJJ.**
- **From Europort, PA3EPD, semi-portable. Is this the future for many of us?**

FTP your antenna pictures to: nitehawk.com/pub/upload/rasmit

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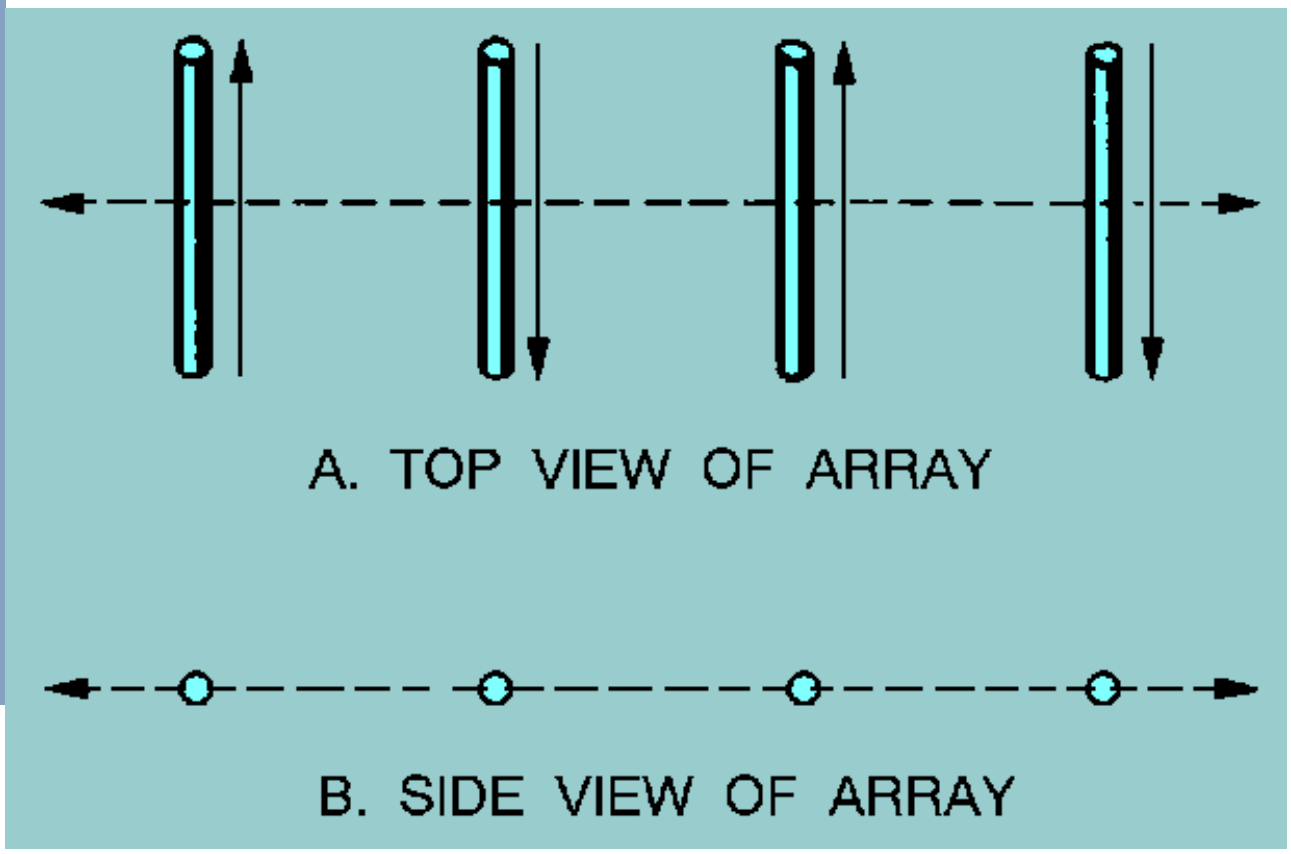
End-Fire Arrays

An end-fire array looks similar to a broadside array. The ladder-like appearance is characteristic of both (fig. 4-28, view A). The currents in the elements of the end-fire array, however, are usually 180 degrees out of phase with each other as indicated by the arrows. The construction of the end-fire array is like that of a ladder lying on its side (elements horizontal). The dipoles in an end-fire array are closer together (1/8-wavelength to 1/4-wavelength spacing) than they are for a broadside array.

Figure 4-28. - Typical end-fire array.

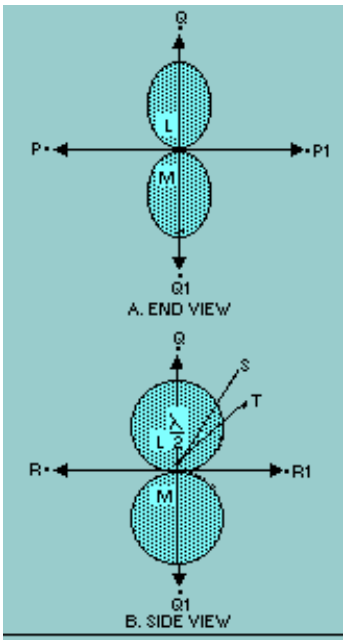


Closer spacing between elements permits compactness of construction. For this reason an end-fire array is preferred to other arrays when high gain or sharp directivity is desired in a confined space. However, the close coupling creates certain disadvantages. Radiation resistance is extremely low, sometimes as low as 10 ohms, making antenna losses greater. The end-fire array is confined to a single frequency. With changes in

climatic or atmospheric conditions, the danger of detuning exists.

RADIATION PATTERN. - The radiation pattern for a pair of parallel half-wave elements fed 180 degrees out of phase is shown in figure 4-29, view A. The elements shown are spaced 1/2 wavelength apart. In practice, smaller spacings are used. Radiation from elements L and M traveling toward point P begins 180 degrees out of phase. Moving the same distance over approximately parallel paths, the respective wavefronts from these elements remain 180 degrees out of phase. In other words, maximum cancellation takes place in the direction of P. The same condition is true for the opposite direction (toward P1). The P to P1 axis is the line of least radiation for the end-fire array.

Figure 4-29. - Parallel elements 180 degrees out of phase.



Consider what happens along the QQ1 axis. Energy radiating from element M toward Q reaches element L in about 1/2 cycle (180 degrees) after it leaves its source. Since element L was fed 180 degrees out of phase with element M, the wavefronts are now in the same phase and are both moving toward Q reinforcing each other. Similar reinforcement occurs along the same axis toward Q1. This simultaneous movement towards Q and Q1 develops a bidirectional pattern. This is not always true in end-fire operation. Another application of the end-fire principle is one in which the elements are spaced 1/4 wavelength apart and phased 90 degrees from each other to produce a unidirectional pattern.

In figure 4-29, view A, elements A and B are perpendicular to the plane represented by the page; therefore, only the ends of the antennas appear. In view B the antennas are rotated a quarter of a circle in space around the QQ1 axis so that they are seen in the plane of the elements themselves. Therefore, the PP1 axis, now perpendicular to the page, is not seen as a line. The RR1 axis, now seen as a line, is perpendicular to the PP1 axis as well as to the QQ1 axis. The end-fire array is directional in this plane also, although not quite as sharply. The reason for the greater broadness of the lobes can be seen by following the path of energy radiating from the midpoint of element B toward point S in view B. This energy passes the A element at one end after traveling slightly more than the perpendicular distance between the dipoles. Energy, therefore, does not combine in exact phase toward point S. Although maximum radiation cannot take place in this direction, energy from the two sources combines closely enough in phase to produce considerable reinforcement. A similar situation exists for wavefronts traveling toward T. However, the wider angle from Q to T produces a greater phase difference and results in a decrease in the strength of the combined wave.

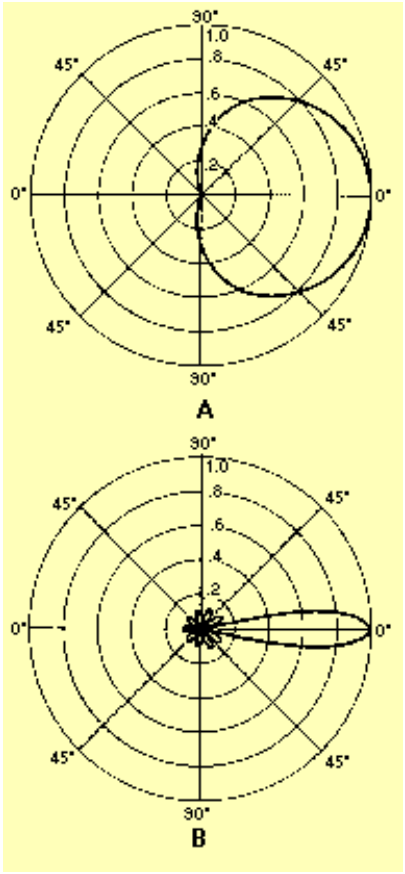
Directivity occurs from either one or both ends of the end-fire array, along the axis of the array, as shown by the broken arrows in figure 4-28, view A; hence, the term *end-fire* is used.

The major lobe or lobes occur along the axis of the array. The pattern is sharper in the plane that is at right angles to the plane containing the elements (figure 4-29, view A). If the elements are not exact half-wave dipoles, operation is not significantly affected. However, because of the required balance of phase relationships and critical feeding, the array must be symmetrical. Folded dipoles, such as the one shown in figure 4-20, view A, are used frequently because the impedance at their terminals is higher. This is an effective way of avoiding excessive antenna losses. Another expedient to reduce losses is the use of tubular elements of wide diameter.

GAIN AND DIRECTIVITY. - In end-fire arrays, directivity increases with the addition of more elements and with spacings approaching the optimum. The directive pattern for a two-element, bidirectional system is illustrated in figure 4-29. View A shows radiation along the array axis in a plane perpendicular to the dipoles, and view B shows radiation along the array axis in the plane of the elements. These patterns were developed with a 180-degree phase difference between the elements. Additional elements introduce small, minor lobes.

With a 90-degree phase difference in the energy fed to a pair of end-fire elements spaced approximately 1/4 wavelength apart, unidirectional radiation can be obtained. The pattern perpendicular to the plane of the two elements is shown in figure 4-30, view A. The pattern shown in view B, taken in the same plane, is for a six-element array with 90-degree phasing between adjacent elements. Since both patterns show relative gain only, the increase in gain produced by the six-element array is not evident. End-fire arrays are the only unidirectional arrays wholly made up of driven elements.

Figure 4-30. - Unidirectional end-fire arrays.



Q.35 What are some disadvantages of the end-fire array? Answer

Q.36 Where does the major lobe in the end-fire array occur? Answer

Q.37 To maintain the required balance of phase relationships and critical feeding, how must the end-fire array be constructed? Answer

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***We may substitute on terminal type, snap versus solder type.**

Properties of Fiber Glass Rods and Tubes

KQ6RH

(C) 1998, 1999, 2000

Ray Jurgens

(Up-Dated 2/25/2000)

Properties of Fiber Glass Rods and Tubes

At this point, I have obtained all of the fiberglass materials that have been using in my antenna projects from Max Gain Systems. These materials are all listed on their web page along with size, weight, color, and price. However, if you are planning to build light weight structures more information is probably required. In particular the bending properties associated with cantilever supports. The lightest weight material is the 1/8" rod, and an 8 ft length is easily bent into a full loop. Thus, this material can only support a small amount of weight at small lengths. The larger diameters of tubing can support increasing amounts of weight for a given deflection. The deflection information is particularly useful in predicting the total bending of a composite spreader and also for determining the amount of tension required in a guy line to re-establish specific position of the spreader.

The tables below give bending information for each of the rods and tubes I have tested. Each size of rod or tubing will bend a specific amount when loaded depending upon the size and length. These materials will also buckle when loaded axially. I concentrated mostly in determining the properties of the materials used as cantilever beams supported rigidly on one end in a horizontal position. Under this condition, the unsupported end will deflect downward due to the gravity load along the entire length of the beam. I measured the unloaded deflection as a function of length for each size material up to and including the 3/4" tubing. I also measured the 1/8" and 1/4" solid rods in the same manner. In general, all of these materials are formed from fiberglass of uniform density and modulus of elasticity, so the bending properties are predictable with a single equation. In general, the deflection increases with the cube of length for a given size and geometry. Further loading at the unsupported end increases the deflection in proportion to the loading, i.e., the lever acts as a linear spring about the rest position. This linear behavior is exhibited so long as the deflection angle is small (less than

about 15 degrees). An 8' span of the 1/8" rod material will deflect more than this angle, and the load no longer pulls perpendicular to the beam. For this reason, the smaller diameter materials are useful only in small lengths (as micro extensions). As an example, the table below shows the properties of the 1/2" tubing. Negative loading implies that the force is applied upward. Normally, I made some attempt to determine the amount of force required to bring the beam into horizontal position at the unsupported end.

grams loading at end of beam	displacement in inches	Comments
-160	0.00	Negative loading at tip
0	6.75	No loading
100	11.00	Positive loading at tip
200	16.50	
300	20.50	

Table 1
Deflection of 8' Cantilever Beam
0.5" OD 0.25 ID Tubing

It is clear from Table 1 that the useful range of loading is probably less than 200 grams for the 8' length. Further loading causes significant droop and such large distortions would be unable to support the wire beams in an acceptable manner. However, one way around this problem is to guy the beam into place. But, it may be more complicated to make a structure such that the beam is guyed from three or four sides. An other alternative is to deliberately bend the beam with a guy under tension from one direction such that this tension is much greater than the loading to be supported. This later procedure works well for making horizontal spreader structures for planar wire beams. The **Reflected M** beams pictured in that section are designed in this manner.

Table 2 below shows the deflection measurements for lengths of the 0.5" tubing for 2', 4' and 8' lengths supported horizontally and loaded at the free end. Note that some deflections were too small to measure, and some others were too large. In general the accuracy o f the deflection measurement is limited to about an 1/8 of an inch, and the gram loading is accurate to about 5%.

grams loading	2' length deflection	4' length deflection	8' length deflection

-160 (pulling up to level)			0.00
-120 (pulling up to level)		0.000	
0 (unloaded)	0.000	0.375	6.75
100 (pulling down)		1.250	11.00
200		1.375	16.50
300	0.313	2.625	20.50
400		3.500	
500	0.500	4.250	

Table 2
Deflection of 0.5" Diameter Tubing Under Loading for Various Lengths

At this point, it is probably best to try to make some sense of these measurements so that the standard equations for beam deflection and loading can be applied. Toward this end, several things are needed, and these include the weight per unit length of the materials, their second moments of inertia, and the modulus of elasticity. We have information on most of these except the latter item which is probably constant for the material. The deflection measurements permit us to get at this number using the deflection equation for uniformly loaded beams. If we let y be the downward deflection, then the equation is

$$y = w * L^4 / (8 * E * I)$$

where w is the weight per unit length (lbs/inch), L is the length (inches), E is the modulus of elasticity (lbs/inch^2) and I is the moment of inertia (inches^4). So, it is easy to see that we can solve for E given all the other parameters. Then, once we know E we can solve for deflections for many other cases. Table 3 is a summary of the data and resulting values of E determined for some of the smaller materials available. The value of the moment of inertia, I, can be calculated for these circular cross sections as

$$I = (pi/4) * (Ro^4 - Ri^4)$$

where Ro is the outside radius and Ri is the inside radius (not diameters). If you use diameters in this equation, you need to divide the result by 16 to get the correct answer. The weights for these materials are located in the MGS home page, but they can be calculated quickly by multiplying the the density of the material by the cross sectional area. This gives the weight per

unit length. The density of these fiber glass materials is about 6.72 x 10^-2 lbs/inch^3 or roughly 1.86 grams/cm^3. The white bicycle whip material in the table is about twice this dense (item # 3).

#	Outside Dia	Inside Dia	Length	y displacement	I	w	E
	(inches)	(inches)	(feet)	(inches)	(inches^4)	(lbs / inch)	(lbs/inch^2)
1	0.125	0.00	2.00	0.750	1.50x10^-5	8.33x10^-4	3.07x10^6
2 black	0.250	0.00	4.00	2.375	2.44x10^-4	3.23x10^-3	3.69x10^6
3 white	0.250	0.00	4.00	1.750	2.44x10^-4	6.43x10^-3	1.00x10^7
4	0.375	9/64	4.00	1.250	1.89x10^-3	6.72x10^-3	3.02x10^6
5	0.500	9/64	8.00	6.750	3.69x10^-3	8.72x10^-3	3.71x10^6
6	0.500	9/64	4.00	0.375	3.69x10^-3	8.72x10^-3	3.70x10^6
7	0.750	0.500	8.00	3.750	1.59x10^-2	1.56x10^-2	2.78x10^6

Table 3
Determination of Modulus of Elasticity from Deflection Measurements

Notice that the value of E is around 3x10^6 psi for all of the black fiber glass samples, but there is some variation that may indicate slight differences in the material. However, using a value near 3x10^6 will provide a good first estimate for engineering purposes, and the variation in the table can be used as an estimate of the expected variance. Note that the white bicycle whip

material is very different. It is twice a dense as the other materials in the table and has a modulus of elasticity that is 3 times as high (about equal to that of aluminum) as the other samples. The E values for the fiber glass are about 1/10 as large as structural steel (for perspective).

The tables that follow are measurements of loading applied to the ends of the horizontal cantilevers. So as long as the displacements are small, the beams respond as linear springs so that the displacement about the quiescent displacement is proportional to the applied load. So we can assume that y = K*W where W is the applied load. The spring constant, K, can be derived from the moment diagram of the loading, however, it may be adequate to know the dependence on the length of the beam is as K ~ L^3/(E*I), so the spring is very stiff for short lengths and gets loose rapidly as the length of the beam increases. So, y ~ L^3 * W / (E * I), and you can find the constant of proportionality for the data in the tables so that you can use W in what ever units you like. Also, note, the 0.5" tubing is given in Table 2 and is not repeated here.

L length in feet	W grams weight applied	Displacement inches
2	0	0.75
2	50	10.00

Table 4
Displacement for 1/8" Diameter Solid Fiber Glass Rod

L length in feet	W grams weight applied	Displacement in inches
2	0	~0
2	100	1.75
2	200	3.00
2	300	4.50
4	0	2.37
4	100	15.50
5' 8"	0	11.50
5' 8"	50	28.00 * beyond linear

Table 5
Displacements for 1/4" Solid Black Fiber Glass

L length in feet	W grams weight applied	Displacement in inches
4	0	1.75
4	100	9.50
4	200	15.25
4	300	19.75
5' 8"	0	6.00
5' 8"	50	20.00
5' 8"	100	24.50

Table 6
Displacements for 1/4" Solid White "Fiber Glass ?"

L length in feet	W grams weight applied	Displacement in inches
2	0	~0
2	100	0.375
2	200	0.625
2	300	0.875
2	400	1.125
2	500	1.500
4	0	1.250
4	100	4.250
4	200	6.375
4	300	8.500
8	0	13.625
8	100	29. * beyond linear limit

Table 7
Displacements for 3/8" Fiber Glass Tubes

L length in feet	W grams weight applied	Displacement in inches
4	0	~0
4	100	0.250
4	200	? got lost
4	300	1.000
8	0	3.750
8	100	5.500
8	200	7.875
8	300	9.875

Table 8
Displacements for 3/4" Fiber Glass Tubes

These tables are useful in determining the displacements of the beams when loading is applied and can be useful for estimating deflections when the supports are at differing angles, such as 45 degrees. In general, one should consider the wire load carefully, and choose materials strong enough to carry the load without serious distortion to the structure. The structures can be made more stable by using nylon monofilament guy lines, however, buckling, and other problems might appear if the guys or wires are too tight. Since the displacements depend upon the length cubed when loads are applied at the tips, the total displacement of compound or telescoping sections can be estimated from the tables above. More information on these topics will be added from time to time.

I have recently spent some time testing kevlar thread in place of the monofilament nylon fish line. This material is very strong and hardly stretches at all. However, this very property can also be a problem in getting the loading distributed equally in the various guy lines. This also means that the lengths must be measured and cut very exactly, and one must know exactly what length to cut. In systems with multiple guy lines, tension on one line affects the others, i.e., all parts of the structural system are coupled. Since these systems present considerable computational difficulty, you can save a lot of grief by including a small turnbuckle in each line. This allows individual adjustment of the tensions. The turnbuckles should be placed at the ends of the guys at the most stable point (for example the guy post). Alternatively, separate guy tie rings can be used for each guy set allowing adjustment of individual guy sets. This second method usually is lighter in weight. You should take care that each guy set is made with

identical lengths even though the actual lengths can depart slightly from those calculated.

The 1/8" material is useful for short extenders for either the 1/4" or 3/8" tubes. I found that terminal lugs can be crimped on to the 1/8" inch rods and grommets inserted to provide a wire guide. The wires can be secured in proper location using a 4" piece of #18 copper wire twisted around the lug and then to either side of the wire.

1/8" Extenders with Crimp Lugs

Return to Main Menu

Flags and Pennants

Receive antennas for limited space with good noise rejection like the Ewe. They are ground independent, too, a big plus!

Here is a compendium of information about these excellent low-noise receive antennas. The Pennant was originated by EA3VY and optimized for 160 meters by K6SE, who first wrote about them on the Top Band Reflector in 1998. It also performs very well on 80 and 40 meters. I built a Flag antenna pointed at Europe and was pleased with the results. I plan to build more of them. K6SE has done a great job in refining and popularizing these antennas. JF1DMQ actually wrote an earlier article about the Flag antenna in November 1995 in a Japanese magazine. His was only 3.3 feet by 16.4 feet long (1 by 5 m). K6SE's 160m optimized versions are 14 by 29 feet (4.3 by 8.8m).

- - Updated 26 Mar 2002 - -

New Pennant Antenna With Remote Termination Control by WA1ION
New Reduced Size and Elevated Flag Antenna Models by VE6WZ

See **QST Magazine**, July 2000, page 34 for K6SE's article:
"Flags, Pennants, and Other Ground-Independent Low-Band Receiving Antennas"

Basic Pennant Description	**More Basic Information**	**More Basic Information**
Size, Pattern, Termination	**Basic Information Q&A**	**G4VGO Reports Good Results**
Elevation Angle, Multi-Antenna Interaction	**Antenna/Mast Interaction**	**Minimizing Interaction With TX Verticals**
TX Vertical Interaction, Transformer Details	**Switching Multi Antennas- Point-Fed Pennant Is Best**	**Multi Antennas- Regular Coax Switches NOT OK**

Performance Compared to Beverages, Phased Vert	Matching Transformer Theory	Final Consensus On Transformer Design?
More Transformer Design, TX Vertical Interaction	Effect Of Termination Resistance On F/B Ratio	Termination Res Critical To Good Noise Rejection
EA3VY Magazine Article, Design Comments	W7IUV's Rotary Flag	Diamond Configuration
Signal Output Is Low But Performance is Good	Importance of F/B Ratio	Termination Res & F/B, More Transformer Notes
More Transformer Theory	More Transformer Theory	More Transformer Theory
More Transformer Theory	More Transformer Theory	Pennant & Flag Arrays
Phasing Loop Arrays	Phasing Loop Arrays	Phasing Loop Arrays
Antenna Comparisons- Very Informative!	Flag Antenna User Report	W8JI Flag Comments
K6SE Comments on Rotary Flag	Delta RX Antenna	JF1DMQ Flag Predecessor
Flag Performance Deterioration	Delta RX Disappointment	Pennant Remote Termination Control
Pennant Working?	Pennant Preamps	Pennant Preamps
Pennant Preamps	W7IUV Compares Low Noise Antennas	Room for 4 Pennants- Part 1
WA1ION & K6SE Q&A	K9AY Comments	Room for 4 Pennants- Part 2

HOME K3KY's DX Toolbar The Low Bands PREV

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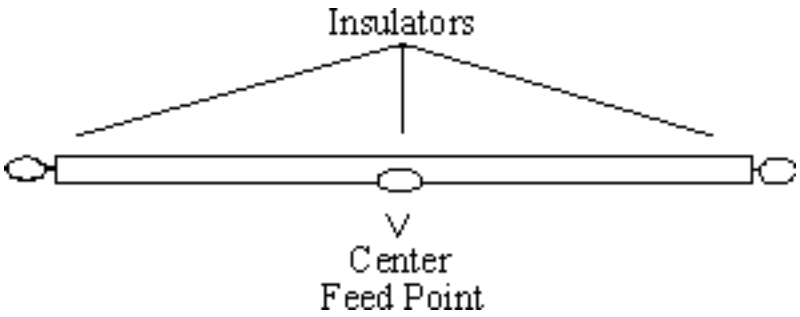
057482

Folded Dipoles

A variation of the dipole is an antenna called a folded dipole. It radiates like a dipole but sort of looks like a **squashed** quad.

Having a folded dipole does **not** mean that you have an antenna that is folded in half and so you obtain an antenna that now takes up half the space of a regular dipole. No, the antenna is still approximately the same length as a regular dipole. It ***is*** however, an antenna that has a wire folded back over itself, hence its name. Below is a picture of a folded dipole.



The starting formula for the folded dipole calculation is the same as a dipole, 468 / Frequency (in MHz). Let's try an example: Design a folded dipole for the 40 meter band. The frequency that is chosen might be 7.15 MHz. Plugging this in to the formula (468 / 7.15) gives a folded dipole with a length of 65.45 feet. When I modeled a dipole on the computer at 30 feet, I came up with a length of 65.47 feet. When I added a second wire to make the folded dipole shown above, I designed the antenna with 1 inch spacing between the two wires. Note that this adds 1 more inch to each of the two antenna wires over that of a single wire dipole. This plus the fact that we are actually turning up the ends of the antenna, means that the horizontal length actually need to be a little shorter to be once again at resonance. The total length came to 64.38 feet, 1.09 feet shorter than the straight dipole. If you use a greater spacing, say 1 foot between the wires the length is 63.1 feet, 2.37 feet shorter. So be sure to shorten the antenna a bit or you'll find yourself operating lower down the band than you expected.

The feed point impedance is also modified by the second wire. Let's say the original dipole was 72 ohms. The step-up for a two wire folded dipole is 4 times which means 4 * 72 = ~288 ohms. (The computer shows 281 ohms on my example, but remember, we reduced the length slightly also.) This step up continues if you add more and more wires. A three wire antenna would provide a step-up of 9, and a four wire antenna provides a step-up of 16.

We can see why this step-up occurs by looking at the power formula P=(I*I) * R, this can be rewritten as R = P / (I*I). If the power to a regular dipole antenna was 100 watts and the current was 1.2 amps, we'd solve for R as R = 100 / (1.2*1.2), which is the same as R = 100 / 1.44, which is 69.44 ohms. In the folded dipole the wires are in parallel, the current must be divided between the two wires. The current in each is half and the total power has not changed, so now the formula is R = 100 / (.6*.6), which is the same as R = 100 / .36, which of course is 277.77 ohms, 4 times the normal dipole antenna.

So now you ask, why would anyone want an antenna with a feed point impedance of 277 ohms, my coax cable is 50 ohms!? Well let's say you wanted to feed the antenna, not with 50 ohm cable but with 300 ohm twin lead? Ah ha, now we have a decent match and a feed line that can also handle a higher SWR with low loss. You'd probably use a tuner (ATU) in the shack to match the 50 ohm radio to the 300 ohm feed line. You could also use the antenna on other bands with the tuner and have an efficient antenna system.

What are the drawbacks to the antenna? Well for one, the currents on each wire will begin to cancel each other out on even multiples of the *cut* frequency, so a 40 meter folded dipole should not be used on 14 MHz. On other bands even though the signal may cancel broad side to the antenna, you'll find that there is actually gain! This occurs about 45 degrees off broad side to the antenna. And this might make for interesting contacts.

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Back to the home page

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Full Wave Delta Loop | 2 Meter Dipole Antenna
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FULL WAVE LOOP ANTENNA

(Plus Halyard Building Instructions)

Are you looking for an inexpensive wire antenna that makes possible HF operation on all bands 10M through to 80M with wide bandwidth?

If yes, you may want to consider putting up this **Delta Loop** like Elizabeth, VE7TLK did on her single residential lot, with the help of amateur radio friends. Antenna apex is up approximately 82' tied to a cedar tree, while the two corners are tied to a shorter tree 25' high and a carport roof.

This Delta Loop is a three-sided antenna suspended high in the air by vertical supports, such as tall evergreen trees. Recommended height is 40 feet or more at highest point, but higher is better. It's one feed line eliminates the need for multiple antennas to cover the HF bands.

Diagram

The Delta Loop's feedpoint is located near one of the bottom corners providing a slight increase in gain and easy access when maintenance is required.

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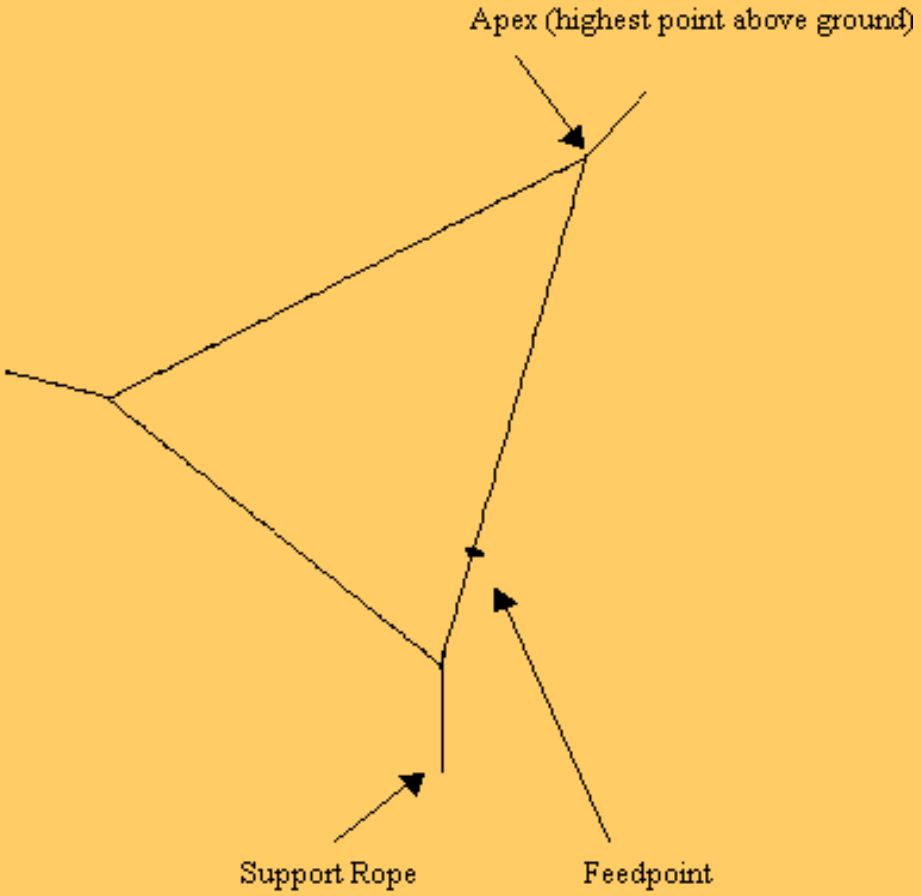

Directional Characteristics

The radiating element or wire position in relation to the ground determines polarization. If the wire is parallel to the ground, it radiates horizontally. If the wire is perpendicular to the ground, it radiates a vertical wave. If the wire is slanted, it radiates waves, which have both horizontal and vertical qualities.

Calculating Length Of Wire Needed For 80 M Loop

$$\text{Length (ft.)} = \frac{1005}{f\ (\text{MHz})}$$

Therefore, length of wire needed for an 80M loop antenna is:

$$\text{Length (ft.)} = \frac{1005}{3.7\text{MHz}} = 272 \text{ feet}$$

Materials Needed For Antenna

- Copper Stranded Wire #12
- 4 ceeramic or glass insulators (apex, two corners, and at feedpoint)
- RG58 or mini RG8 Coax (50 ohm) Calculate length from feed line to radio room plus a little extra.
- PL259 Connector (UHF male) - for end of coax which connects to radio.
- Support rope for corners and apex.
- Roll of coax seal to wrap.
- Polyester braided rope (3/8") needed for halyard and support ropes.

Step By Step Instructions

1. Draw to scale diagram of antenna and all supports with dimensions on lot.

2. Measure and cut wire length for antenna allowing 1' extra for securing insulators. Careful not to bend wire.

3. Lay the wire on the ground so that the sides can be measured and the insulators fixed to the apex and corners. To fix insulator cut a short piece of wire and twist it around the antenna wire for about 4" on one side and then cross over insulator end and repeat 4" up the other wire. This adds strength to the corners and apex ands keeps insulators in place when raising antenna.
4. Cut three pieces of support rope to clear and free antenna corners and apex from trees/buildings, etc. Connect ropes to the three insulators.

5. Feedpoint connection ¼ wavelength from bottom corner. (See diagram) With a sharp knife, carefully strip back (not cut) 3" of the exterior black jacket from one end of the 50 ohm coax being careful not to score the braid underneath. Pull the braid gently a part in one place and pull out the inner dielectric containing the center conductor. The coax is now split in two. Expose the center conductor by carefully cutting off 1" of the inner dielectric which is covering the center conductor.

6. Each wire end is then threaded through the closest insulator hole and then tightly twisted around antenna wire on that end. This is also repeated on the other end of the same insulator using the other antenna wire. Wrap prepared coax end around the middle of the insulator and secure with a clamp.

7. Once this is done and the materials have been cleaned

well, solder one of the #12 gauge copper wires to braid and the other to the center conductor. To avoid water and contamination in the feed line, the antenna end of the feed line must be adequately covered with coax seal . Remember, the feed line connection at the center insulator should not be done until after the antenna wire has been tied securely to the insulator.

8. Construct a 1:1 broadband balun. Wind or coil a length of coaxial feed line for 6-8 turns near point of connection and secure with electrical tape. Lengths are not critical. Diameter of coils approximately 8".

9. Pull the antenna wire carefully up in the air by pulling down on the rope end which will eventually be tied to the limb of the tree. Once desired height of highest corner or apex has been reached, tie rope to limb near ground and within reach by a ladder.

10. One of the two remaining insulators will have its rope now tied to the second tree (or other support) which also has a collar but no pulley. This insulator and the previous one are counterweighted (halyard) and allowed to move freely.

11. The delta loop is now ready to be raised using the pulley on the halyard at the top of the tree or pole. Once the apex is fastened at the correct height proceed to fasten securely the corner ropes to their tie supports.

12. Install the PL259 (UHF male) Connector on end of coax feedline in radio room. An adapter is available if using the smaller RG-58 size coax. Connect to transceiver and test for SWR.

Halyard Recommended

Introduces slack so that during high wind conditions the wire loop can move, thus reducing stress on wire avoiding breakage. See directions for making halyard.

Materials And Directions

1. Bucket (with several holes on bottom) with handle plus counterweight weighted with bricks or sand and rocks.

2. One pulley large enough for line to run through freely keeping in mind that rope can swell when wet. Buy a fast-eyed brass or bronze marine type pulley.

3. Polyester braided rope (3/8") needed to go up and down

full length of tree trunk from collar to ground. Allow for extra to tie around lower limb and to raise antenna without running out of rope.

4. Make a trunk collar to protect bark of tree. This is done by threading a short piece of rope through an old hose or other soft tubing, the diameter of the trunk, where the collar will be positioned up high on the tree. Thread the pulley onto collar rope where the two ends of collar will meet together and tie securely. Only an experienced tree climber, with safety climbing equipment, should do the climbing.

5. While climber is up on the tree have him take one end of the long rope and feed it through the pulley. As he comes down he can pull that end down with him to ground level. Do not cut rope yet until after the antenna has been raised up, otherwise you may find yourself with too short a rope.

6. One end of rope will be tied to the weighted bucket handle and the other end tied tightly to a low limb within your reach. Bucket should be suspended about six feet from ground floating freely. The antenna wire at this point is still lying on the ground.

7. Just before the antenna wire is ready to be raised, tie the appropriate support short rope to the insulator at the apex of the antenna to the halyard rope's mid-point. This rope enables the wire to position itself away from tree branches.

8. Raise the apex of the antenna by pulling on the rope without the bucket tied to it to the desired height. Then tie rope securely to one of the lower limbs within reach.

Grounding Antenna & Radio

Materials Needed

- 8 Feet Copper Ground Rods (1 to 3) five to six feet apart (?)
- Brass clamps (for each rod)
- Belden grounding braid or 2" copper ribbon
- Good Power Bar with filter and surge protector
- Copper Bar attached to back of station table-top

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G5RV Multi-Band Antenna by Louis Varney,

CEng, MIEE, AIL, G5RV

***** taken from (RADIO COMMUNICATIONS, JULY 1984)

THE G5RV ANTENNA, with its special feeder arrangement, is a multiband centre-fed antenna capable of very efficient operation on all hf bands from 3.5 to 28mhz, specifically designed with dimensions which allow it to be installed in gardens which accommodate a reasonably-straight run of about 102ft (31.1m) for the "flat-top". However, because the most useful radiation from a horizontal or inverted-V resonant antenna takes place from the center two-thirds of its total length, up to one-sixth of this total length at each end of the antenna may be dropped vertically, semi-vertically, or bent at some convenient angle to the main body of the antenna without significant loss of effective radiation efficiency. For installation in a very limited space, the dimensions of both the "flat-top" and the matching section can be divided be a factor of two to make the half-size G5RV, which is a very efficient antenna from 7 to 28 mhz. The full-size G5RV will also function on 1.8mhz band if the station end of the feeder (either balanced or coaxial-type) is strapped and fed by a suitable antenna tuner using a good earth connection or a counterpoise wire. Similarly, the half-size version may be used thus on 3.5 and 1.8 mhz bands.

In contradistinction to multiband antennas in general, the full size G5RV antenna was not designed as a half-wave dipole on the lowest frequency of operation, but as a 1 1/2 wave centre-fed long-wire antenna on 14mhz, where the 34ft (10.36m) open-wire matching section functions as a 1:1 impedance transformer, enabling the 75ohm twinlead or 50/80ohm coaxial cable feeder to "see" a close impedance match on that band with a consequently low vswr on the feeder. However,on all the other hf bands the function of this section is to act as a " make-up" section to accommodate that part of the standing-wave (current and voltage components) which, on certain of the operating frequencies, cannot be completely accommodated on the "flat-top" (or inverted-V)radiation portion. The design centre frequency for the full-size version is 14,150khz, and the dimensions of 102ft (31.1m) is derived from the formula for long-wire antennas which is:

$$\text{Length (ft)} = \frac{492\,(n - 0{\cdot}05)}{f\text{MHz}} = \frac{492 \times 2{\cdot}95}{14{\cdot}15} = 102{\cdot}57\text{ft}\ (31{\cdot}27\text{m})$$

where n= number of half-wavelengths of the wire (flat-top).

In practice, since the whole system will be brought to resonance by the use of an antenna tuner, the antenna is cut to 102ft (31.1m).

As it does not make use of traps or ferrite beads, the "dipole" portion becomes progressively longer in electrical length with increasing frequency.

This effect confer certain advantages over a trap or ferrite-bead loaded dipole because, with increasing electrical length, the major lobes of the vertical component of the polar diagram tend to be lowered as the operating frequency is increased. Thus, from 14mhz up, most of the energy radiated in the vertical plane is at angles suitable for dx working. Furthermore, the polar diagram changes with increasing frequency from a typical half-wave dipole pattern at 3.5mhz

and a 2 1/2 wave in-phase pattern at 7 and 10mhz to that of a "long-wire" antenna at 14, 18, 21, 24 and 28mhz.Figure 1.

Although the impedance match for 75 ohm twinlead or 80 ohm coaxial cable at the base of the matching-section is very good at 14mhz, and even the use of 50 ohm coax cable results in only about 1.8:1 vswr on this band, the use of a suitable antenna tuner is necessary on all the other hf bands because, on those bands, the antenna plus the matching-section will present a reactive load to the feeder. thus the use of the correct type of antenna tuner (unbalanced input to balanced output if twin-wire feeder is used, or unbalanced to unbalanced if coaxial feeder is used) is essential in order to ensure the maximum transfer of power to the antenna from a typical transceiver having a 50 ohm coaxial (unbalanced) output. Also to satisfy the stringent load conditions demanded by such modern equipment employing an alc system which "senses" the vswr condition presented to the solidstate transmitter output stage so as to protect it from damage which could be caused by a reactive load having a vswr of more than about 2:1.

Figure 2

The above reasoning does not apply to the use of the fullsize G5RV antenna on 1.8mhz, or to the use of the half-size version on 3.5 and 1.8mhz. In these cases the station end of the feeder conductors should be "strapped" and the system tuned to resonance by a suitable series-connected inductance and capacitance circuit connected to a good earth or counterpoise wire. Alternately, an "unbalanced-to-unbalanced" type of antenna tuner such as a "T" or "L" matching circuit can be used. Under these conditions the "flat-top" (or inverted-V) portion of the antenna plus the matching section and feeder function as a "Marconi" or "T" antenna, with most of the effective radiation taking place from the vertical, or near vertical, portion of the system; the "flat-top" acting as a top-capacitance loading element. However, with the system fed as described above, very effective radiation on these two bands is obtainable even when the "flat-top" is as low as 25ft (7.6m) above ground.

Theory of Operation

The general theory of operation has been explained above; the detailed theory of operation on each band from 3.5 to 28mhz follows, aided by figures showing the current standing wave conditions on the "flat-top" and the matching (or make-up) section. The relevant theorical horizontal plane polar diagrams for each band may be found in any specialized antenna handbooks. However, it must be borne in mind that: (a) the polar diagrams generally shown in two dimensional form are, in fact, three dimensional (ie solid) figures around the plane of the antenna; and (b) all theoretical polar diagrams are modified by reflection and absorption effects of near-by conducting objects such as wire fences, metal house guttering, overhead electric power and telephone wires, house electric wiring system, house plumbing systems, metal masts and guy wires, and large trees. Also the local earth conductivity will materially affect the actual polar radiation pattern produced by an antenna. Theoretical polar diagrams are based on the assumptions that an antenna is supported in "free space" above a perfectly conducting ground. Such conditions are obviously impossible of attainment in the case of typical amateur installations. What this means in practice is that the reader should not be surprised if any particular antenna in a typical amateur location produces contacts in directions where a null is indicated in the theoretical polar diagram and perhaps not such effective radiation in the directions of the major lobes as theory would indicate.Figure 3

3.5Mhz. On this band each half of the "flat-top" plus about 17ft (5.18m) of each leg on the matching-section forms a fore-shortened or slightly folded up half-wave dipole. The remainder of the matching-section acts as an unwanted but unavoidable reactance between the electrical centre of the dipole and the feeder to the antenna tuner. The polar diagram is effectively that of a half-wave antenna. See figure 1.

7Mhz. The "flat-top" plus 16ft (4.87m) of the matching section now functions as a partially-folded-up "two half-wave in phase" antenna producing a polar diagram with a somewhat sharper lobe pattern than a half-wave dipole due to its colinear characteristics. Again, the matching to a 75 ohm twinlead or 50/80 ohm coaxial feeder at the base of the matching section is degraded somewhat by the unwanted reactance of the lower half of the matching section but, despite this, by using a suitable antenna tuner the system loads well and radiates very effectively on this band. See figure 2.

10Mhz. On this band the antenna functions as a two half-wave in-phase colinear array, producing a polar diagram virtually the same as on 7mhz. A reactive load is presented to the feeder at the base of the matching section but, as for 7mhz, the performance is very effective. See figure 3.

14Mhz. At this frequency the conditions are ideal. The "flat-top" forms a three-half-wave long centre-fed antenna which produces a multi-lobe polar diagram with most of its radiated energy in the vertical plane at an angle of about 14 degrees, which is very effective for dx working. Since the radiation resistance at the centre of a three-half-wave long-wire antenna supported at a height of half-wave above ground of average conductivity is about 90 ohm, and the 34ft (10.36m) matching section now functions as a 1:1 impedance transformer, a feeder of anything between 75 and 80 ohm characteristic impedance will "see" a non-reactive (ie resistive) load of about this value at the base of the matching section, so that the vswr on the feeder will be very nearly 1:1. Even the use of 50 ohm coaxial feeder will result in a vswr of only about 1.8:1. It is here assumed that 34ft (10.36m) is a reasonable average antenna height in amateur installations. See figure 4.

18Mhz. The antenna functions as two full-wave antennas fed in phase; combining the broadside gain of a two-element colinear array with somewhat lower zenith angle radiation than a half-wave dipole due to its long-wire characteristic. See figure 5

21Mhz. On this band the antenna works as a "long-wire" of five half-waves, producing a multilobe polar diagram with very effective low zenith angle radiation. Although a high resistive load is presented to the feeder at the base of the make-up section, the system loads very well when used in conjunction with a suitable antenna tuner and radiates very effectively for dx contacts. See figure 6.

24Mhz. The antenna again functions effectively as a five-half-wave "long-wire" but, because of the shift in the positions of the current anti-nodes on the flat-top and the matching section, as may be seen from figure 7, the matching

or "make-up" section now presents a much lower resistive load condition to the feeder connected to its lower end than it does on 21mhz. Again, the polar diagram is multilobed with low zenith angle radiation.

28Mhz. On this band, the antenna functions as two "long-wire" antenna, each of three half-waves, fed in-phase. The polar diagram is similar to that of a three half-wave "long-wire" but with even more gain over a half-wave dipole due to the colinear effect obtained by feeding two three-half-wave antennas, in line and in close proximity, in-phase. See figure 8.

Construction

The Antenna

The dimensions of the antenna and its matching section are shown in Figure 9. The "flat-top" should, if possible, be horizontal and run in a straight line, and should be erected as high as possible above ground. In describing the theory of operation, it has been assumed that it is generally possible to erect the antenna at an average height of about 34ft (10.36m), which happens to be the optimum radiation efficiency on 1.8, 3.5 and 7mhz for any horizontal type antenna, in practice few amateurs can install masts of the optimum height of half a wavelength at 3.5 or 7mhz, and certainly not at 1.8mhz.

If, due to limited space available, or to the shape of the garden, it is not possible to accommodate the 102ft (31.1m) top in a straight line, up to about 10ft (3m) of the antenna wire at each end may be allowed to hang vertically or at some convenient angle, or be bent in a horizontal plane, with little practical effect upon performance. This is because, for any resonant dipole antenna, most of the effective radiation takes place from the centre two-thirds of its length where the current antinodes are situated. Near to each end of such an antenna, the amplitude of the current standing wave falls rapidly to zero at the outer extremities; consequently, the effective radiation from these parts of the antenna is minimal.

The antenna may also be used in the form of an inverted-V. However, it should be borne in mind that, for such a configuration to radiate at maximum efficiency, the included angle at the apex of the V should not be less than about 120 degrees. The use of 14awg enameled copper wire is recommended for the flat-top or V, although thinner gauges such as 16 or even 18awg can be used.

The Matching Section

This should be, preferably, of open-wire feeder construction for minimum loss. Since this section always carries a standing-wave of current (and voltage) its actual impedance is unimportant. A typical, and very satisfactory, form of construction is shown in figure 10. The feeder spreaders may be made of any high-grade plastic strips or tubing; the clear plastic tubing sold for beer or wine siphoning is ideal.

If it is desired to use 300 ohm ribbon type feeder for this section, it is strongly recommended that the type with "windows" be used because of its much lower loss than that with solid insulation throughout its length, and its relative freedom from the "detuning" effect caused by rain or snow. If this type of feeder is used for the matching section, allowance must be made for its velocity factor (vf) in calculating the mechanical length required to resonate as a half-

wave section electrically at 14.15mhz. Since the vf of standard 300 ohm ribbon feeder is .82, the mechanical length should be 28ft (8.5m). However, if 300 ohm ribbon with "windows" is used, its vf will be almost that of open-wire feeder, say .90, so its mechanical length should be 30.6ft (9.3m). This section should hang vertically from the centre of the antenna for at least 20ft (6.1m) or more if possible. It can then be bent and tied off to a suitable post with a length of nylon or terylene cord so as to be supported at above head-height to the point where, supported by a second post, its lower end is connected to the feeder.

The Feeder

The antenna can be fed by any convenient type of feeder provided always that a suitable type of antenna tuner is used. In the original article describing the G5RV antenna, published in the , then, RSGB bulletin November 1966, it was suggested that if coaxial cable feeder was used, a balun might be employed to provide the necessary unbalanced-to-balanced transformation at the base of the matching section. This was because the antenna and its matching section constitute a balanced system, whereas a coaxial cable is an unbalanced type of feeder. However, later experiments and a better understanding of the theory of operation of the balun indicated that such a device was unsuitable because of the highly reactive load it would "see" at the base of the matching or "make-up" section on most hf bands.

It is now known that if a balun is connected to a reactive load presenting a vswr of more than about 2:1, its internal losses increase, resulting in heating of the windings and saturation of its core (if used). In extreme cases, with relatively high power operation, the heat generated due to the power dissipated in the device can cause it to burn out. However, the main reason for not employing a balun in the case of the G5RV antenna is that, unlike un antenna tuner which employs a tuned circuit, the balun cannot compensate for the reactive load condition presented to it by the antenna on most of the hf bands, whereas a suitable type of antenna tuner can do this most effectively and efficiently.

Recent experiments by the author to determine the importance or otherwise of "unbalance" effects caused by the direct connection of a coaxial feeder to the base of the matching section had a rather surprising result. They proved that, in fact, the hf currents measured at the junction of the inner conductor or the coaxial cable with one side of the (balanced) matching section and at the junction of the outer coaxial conductor (the shield) with the other side of this section are virtually identical on all bands up to 28mhz, where a slight but inconsequential difference in these currents has been observed. There is, therefore, no need to provide an unbalanced-to-balanced device at this junction when using coaxial feeder.

However, the use of an unbalanced-to-unbalanced type of antenna tuner between the coaxial output of a modern transmitter (or transceiver) and the coaxial feeder is essential because of the reactive condition presented at the station end of this feeder which, on all but the 14mhz band, will have a fairly high to high vswr on it. This vswr, however, will result in insignificant losses on a good-quality coaxial feeder of reasonable length; say, up to about 70ft (21.3m). Because it will, inevitably, have standing waves on it, the actual characteristic impedance of the coaxial cable is unimportant, so that either 50 ohm or 80 ohm type can be used.

Another very convenient type of feeder that may be used is 75 ohm twinlead. However, because of the relatively high loss in this type of feeder at frequencies above about 7mhz, especially when it has a high vswr on it, it is recommended that not more than about 50 to 60ft (15.2 to 18.3m) of this type feeder be used between the base of the matching section and the antenna tuner. Unfortunately the 75 ohm twinlead in the UK is the receiver type; the much less lossy transmitter type is available in the USA. By far the most efficient feeder is the "open wire" type. A suitable length of such feeder can be constructed in exactly the same way as that described for the open-wire matching section. If this form of feeder is employed, almost any convenient length may be used from the centre of the antenna right to the antenna tuner (balanced) output terminals. In this case, of course, the matching section becomes an integral part of the feeder. A particularly convenient length of open-wire feeder is 84ft (25.6m), because such a length permits parallel tuning of the

antenna tuner circuit on all bands from 3.5 to 28mhz with conveniently located coil taps in the antenna tuner coils for each band, or, where the alternative form of antenna tuner employing a three-gang 500pf/section variable coupling capacitor is used the optimum loading condition can be achieved for each band. However, this is not a rigid feeder length requirement and almost any length that is mechanically convenient may be used. Since this type of feeder will always carry a standing wave, its characteristic impedance is unimportant, and sharp bends, if necessary, may be used without detriment to its efficiency. It is only when this type of feeder is correctly terminated by a resistive load equal to its characteristic impedance that such bends must be avoided.

Coaxial cable hf choke

Under certain conditions, either due to the inherent "unbalanced-to-balanced" effect caused by the direct connection of a coaxial feeder to the base of the (balanced) matching section, or to pick-up of energy radiated by the antenna, a current may flow on the outside of the coaxial outer conductor. This effect may be considerably reduced, or eliminated, by winding the coaxial cable feeder into a coil of 8 to 10 turns about 6in in diameter immediately below the point of connection of the coaxial cable to the base of the matching section. the turns may be taped together or secured by nylon cord.

It is important, of course, that the junction of the coaxial cable to the matching section be made thoroughly water-proof by any of the accepted methods; binding with several layers of plastic insulating tape or self-amalgamating tape and then applying two or three coats of polyurethane varnish, or totally enclosing the end of the coaxial cable and the connections to the base of the matching section in a sealant such as epoxy resin.

LOW REACTANCE MULTIBAND CENTER FED WIRES

An antenna system is more easily interfaced to a radio when the input reactance at the feedline terminals is low or close to series resonance. If a center fed dipole is fed with a length of low loss balanced transmission line, the input reactance is a function of both transmission line length and antenna length, and to reduce the reactance at selected frequencies there are optimum lengths of dipole and feedline. The G5RV antenna design is an example of selecting these lengths to produce near-series resonances at amateur operating frequencies from 3.5-28MHz. There are other combinations of feedline length and dipole length which reduce the overall reactance as shown below. The red regions of Figure 1 have low RMS reactance over the frequencies shown to the right of the figure.

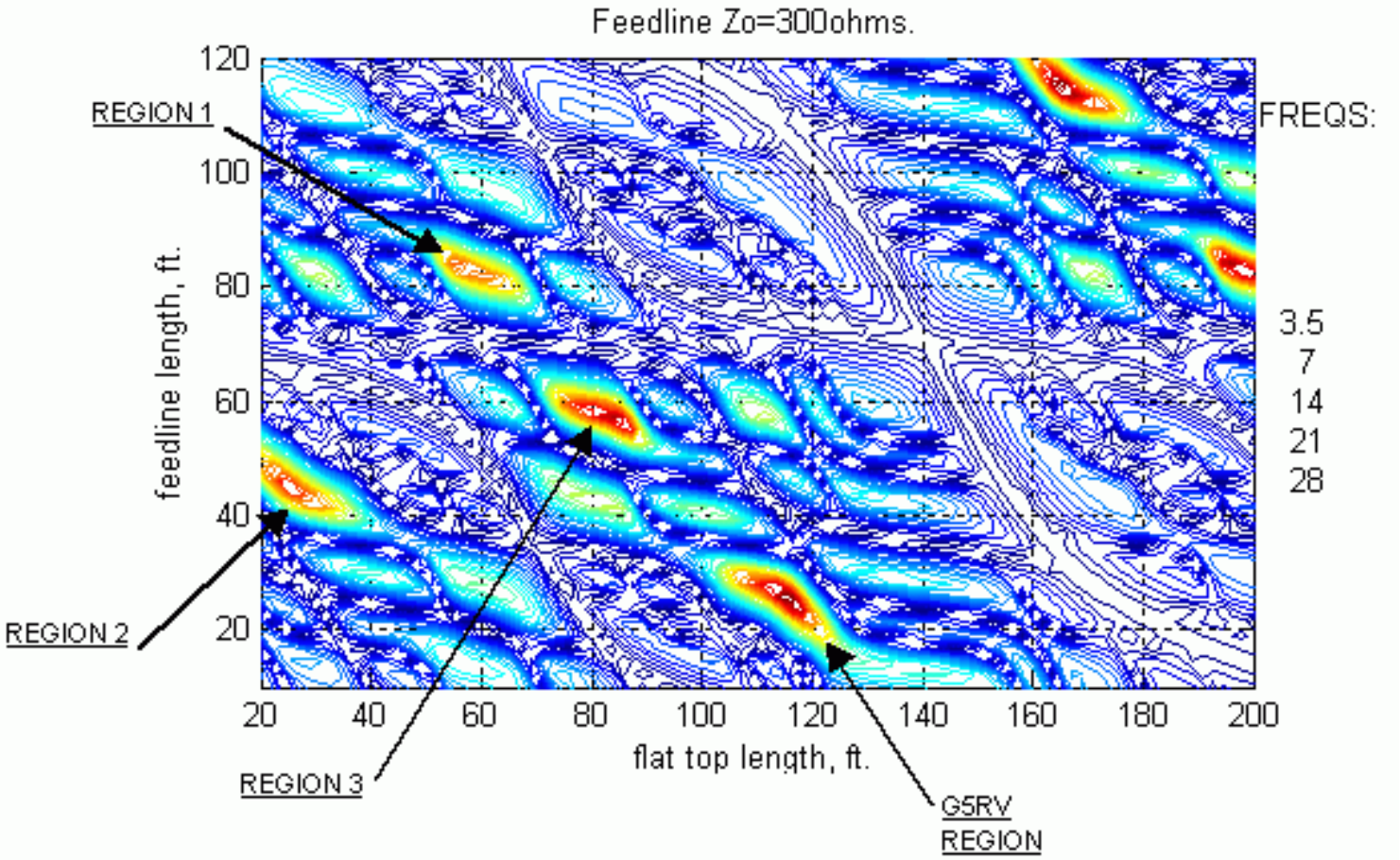

Figure 1. Regions of Low RMS Input Reactance at Amateur Bands.

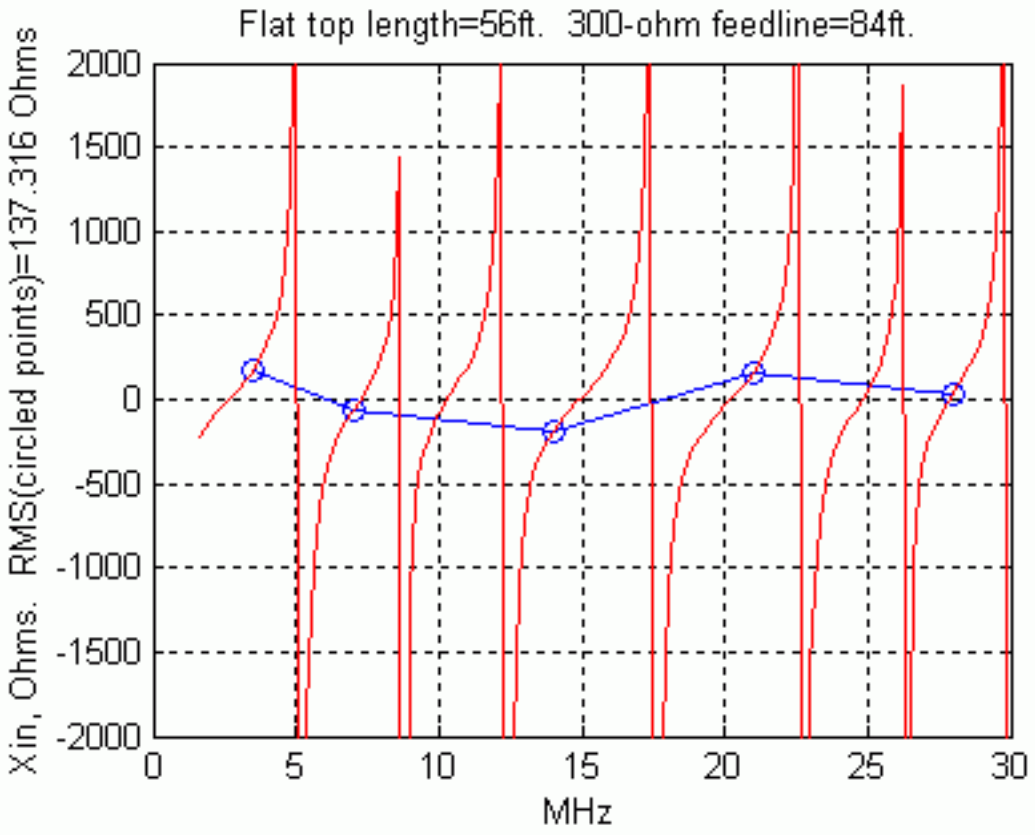

Figure 2. Input Reactance of Region 1.

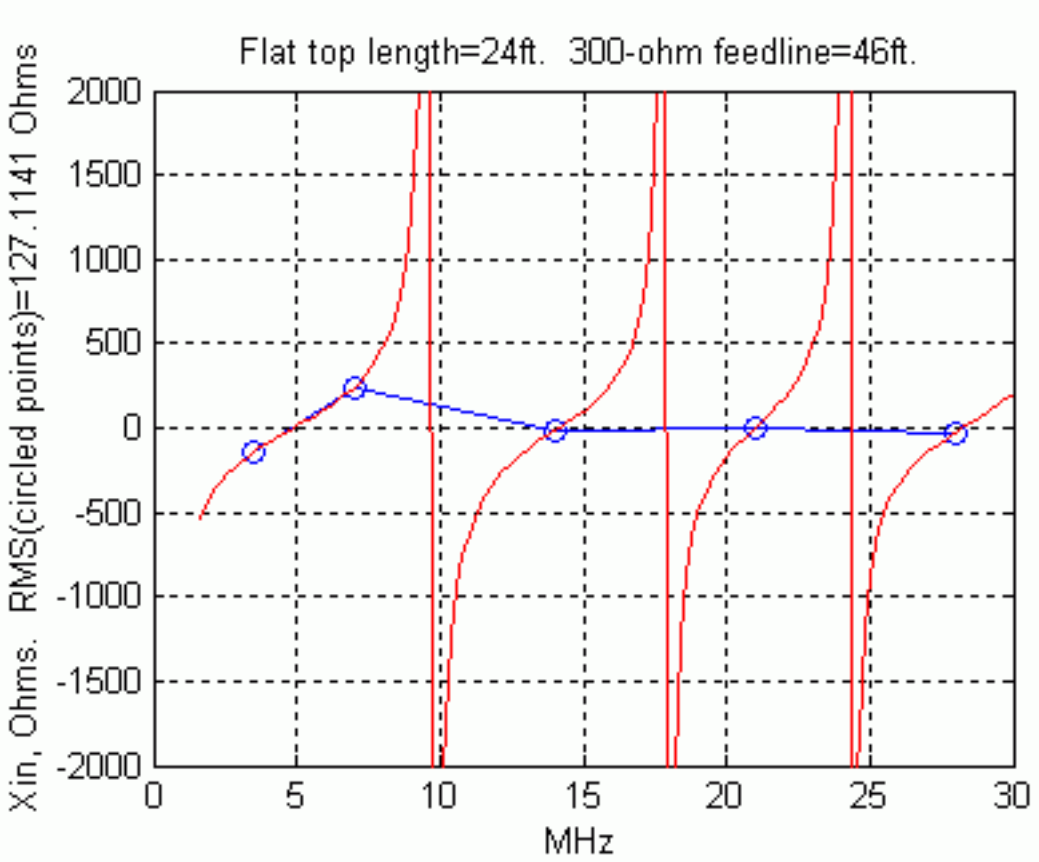

Figure 3. Input Reactance of Region 2.

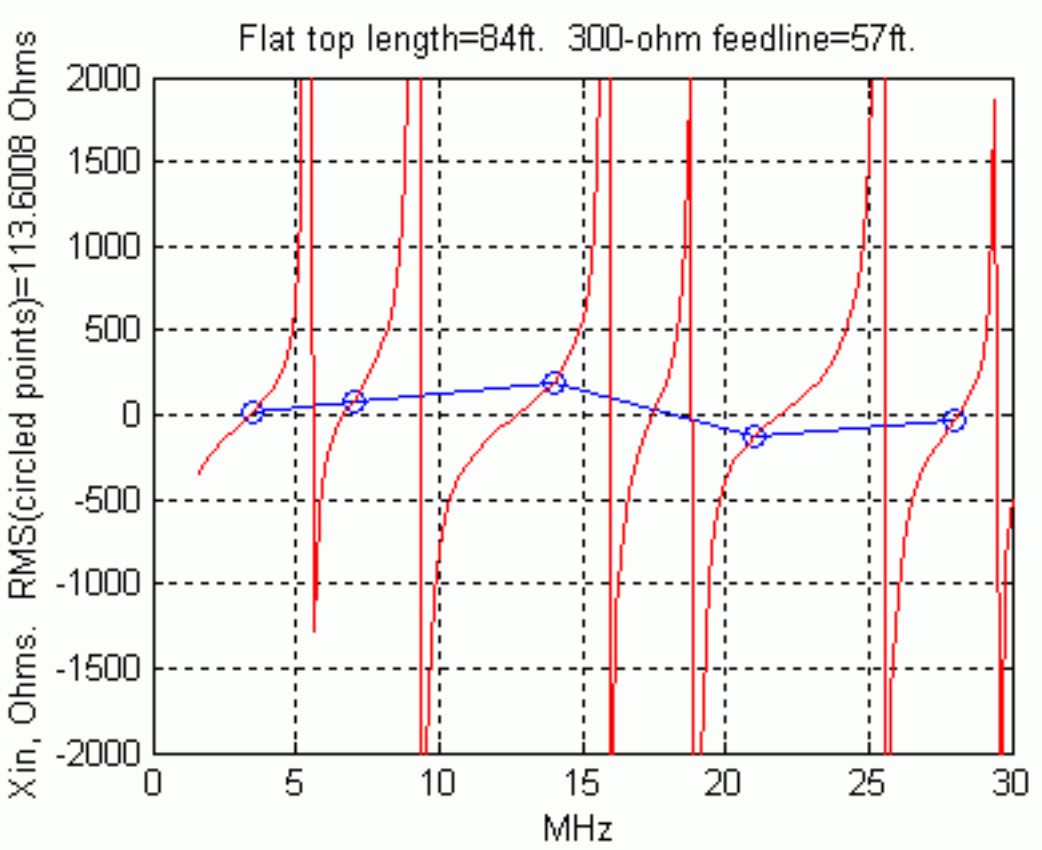

Figure 4. Input Reactance of Region 3.

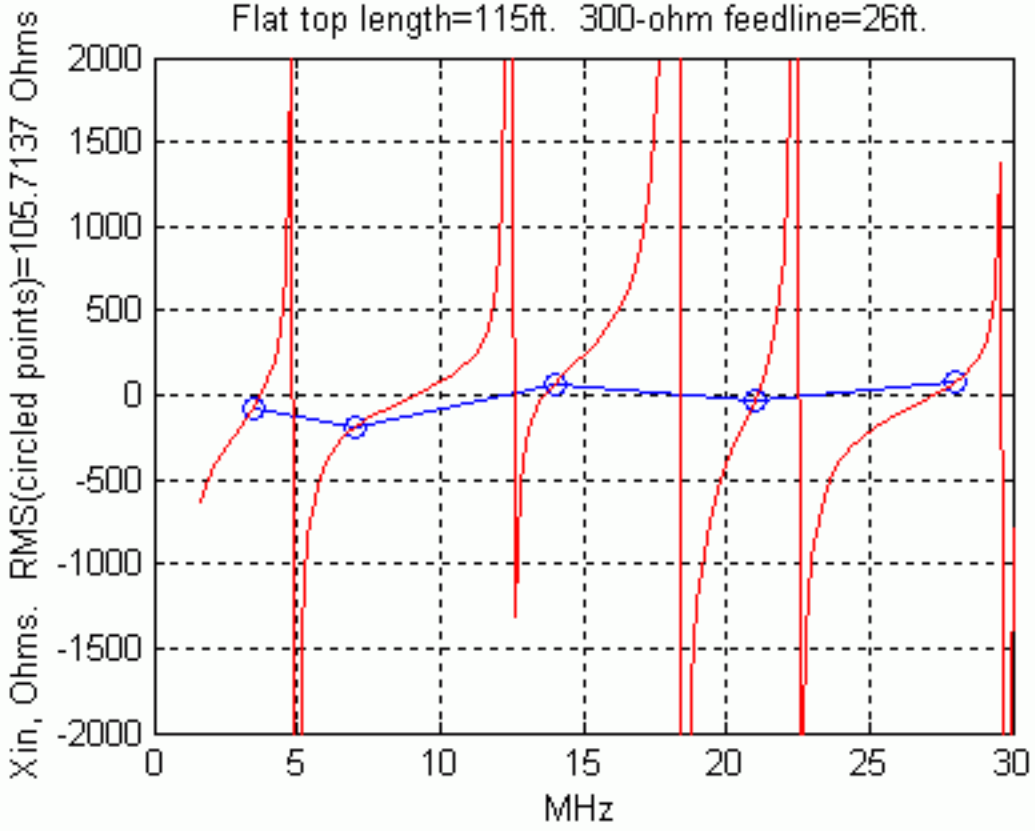

Figure 5. Input Reactance in G5RV Region.

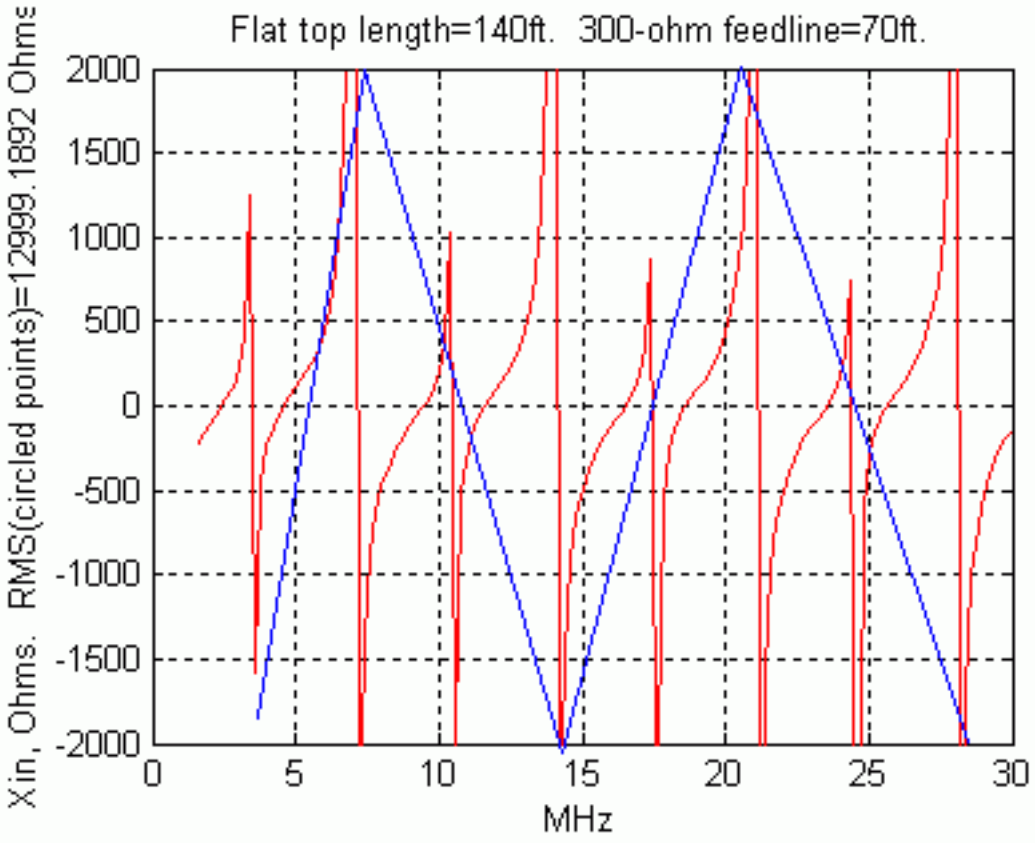

Figure 6. A Bad Choice of Lengths.

Grounding is key to good reception

From: jpd@space.mit.edu (John Doty)
Newsgroups: rec.radio.shortwave
Subject: Grounding is key to good reception, was: Experience w/NRD-535
Date: 16 Feb 1995 16:36:40 GMT

In article <825_9502140342@tor250.org> Larry.Picard@odxabbs.tor250.org (Larry Picard) writes:

> In your recent post you advised that coax should be grounded at two sites, first at the antenna and then just before entering the house. Is there an advantage in grounding at more than these sites?

With grounds the most common experience is "the more the merrier". As you add more, however, you usually reach a diminishing returns (no pun intended) situation where there is no *observable* improvement: that's usually a good place to stop. There are also exceptional circumstances where grounding increases noise problems, but these, in my experience, are much rarer than the pundits who preach against "ground loops" seem to think.

Even a semi-quantitative theoretical treatment of grounding in oversimplified situations requires heavy math at RF. Experimentation is thus required even if one has done elaborate calculations. It's often easier to use the theory as a guide to what to try, and then experiment.

> I would also assume that the antenna is grounded when it is connected to the receiver as the outer braid of the coax is in continuity with the receiver chassis.

What's ground? If connect the shield of my coax (which is grounded outside) to the antenna input of my R8, I hear lots of junk, indicating that there is an RF voltage difference between the coax shield and the R8 chassis. Last night this measured about S5.5, which is about -93 dBm (preamp off, 6KHz bandwidth). That's a lot of noise: it was 18 dB above my antenna's "noise floor", and 26 dB above the receiver's noise floor.

This sort of disagreement about ground potential is characteristic of electrically noisy environments. The receiver will, of course, respond to any voltage input that differs from its chassis ground. The antenna, on the other hand, is in a very different environment, and will have its own idea of what ground potential is. If you want to avoid noise pickup, you need to deliver a signal, referenced at the antenna to whatever its ground potential is, in such a way that when it arrives at the receiver, the reference potential is now the

receiver's chassis potential.

Coaxial cable represents one way to do this. Coax has two key properties:

1. The voltage between the inner conductor and the shield depends only on the state of the electromagnetic field within the shield.

2. The shield prevents the external electromagnetic field from influencing the internal electromagnetic field (but watch out at the ends of the cable!).

So, it's easy, right? Run coax from the antenna to the receiver. Ground at the antenna end will be whatever the antenna thinks it is, while ground at the receiver end will be whatever the receiver thinks it is. The antenna will produce the appropriate voltage difference at the input side, and the receiver will see that voltage difference uncontaminated by external fields, according to the properties given above.

Unfortunately, it doesn't quite work that way. It's all true as far as it goes, but it neglects the fact that the coax can also guide noise from your house to your antenna, where it can couple back into the cable and into your receiver. To see how this works, let me first describe how this noise gets around.

The noise I'm talking about here is more properly called "broadband electromagnetic interference" (EMI). It's made by computers, lamp dimmers, televisions, motors and other modern gadgets. I have all these things. In many cases, I can't get them turned off, because it would provoke intrafamilal rebellion. However, even when I turn them off, the noise in the house doesn't go down very much, because my neighbors all have them too. In any case, one of the worst offenders is my computer, which is such a handy radio companion I'm not about to turn *it* off.

Some of this noise is radiated, but the more troublesome component of this is conducted noise that follows utility wires. Any sort of cable supports a "common mode" of electromagnetic energy transport in which all of the conductors in the cable are at the some potential, but that potential differs from the potential of other nearby conductors ("ground"). The noise sources of concern generate common mode waves on power, telephone, and CATV cables which then distribute these waves around your neighborhood. They also generate "differential" mode waves, but simple filters can block these so they aren't normally a problem.

So, let's say you have a longwire antenna attached to a coaxial cable through an MLB ("Magnetic Longwire Balun" [sic]). Suppose your next door neighbor turns on a dimmer switch. The resulting RF interference travels out his power lines, in through yours, through your receiver's power cord to its chassis, and out your coaxial cable to your MLB. Now on coax, a common mode wave is associated with a current on the shield only, while the mode we want the signal to be in, the "differential" mode, has equal but opposite currents flowing on shield and inner conductor. The MLB works by coupling energy from a current flowing between the antenna wire and the coax shield into into the differential mode. But wait a second: the current from the antenna flows on the coax shield just like the common mode current

does. Does this mean that the antenna mode is contaminated with the noise from your neighbor's dimmer?

The answer is a resounding (and unpleasant) yes! The way wire receiving antennas work is by first moving energy from free space into a common mode moving along the antenna wire, and then picking some of that off and coupling it into a mode on the feedline. In this case, the common mode current moving along the antenna wire flows into the common mode of the coax, and vice versa. The coax is not just feedline: it's an intimate part of the antenna! Furthermore, as we've seen, it's connected back through your electrical wiring to your neighbor's dimmer switch. You have a circuitous but electrically direct connection to this infernal noise source. No wonder it's such a nuisance!

The solution is to somehow isolate the antenna from the common mode currents on the feedline. One common way to do this is with a balanced "dipole" antenna. Instead of connecting the feedline to the wire at the end, connect it to the middle. Now the antenna current can flow from one side of the antenna to the other, without having to involve the coax shield. Unfortunately, removing the necessity of having the coax be part of the antenna doesn't automatically isolate it: a coax-fed dipole is often only slightly quieter than an end-fed longwire. A "balun", a device which blocks common mode currents from the feedline, is often employed. This can improve the situation considerably. Note that this is not the same device as the miscalled "Magnetic Longwire Balun".

Another way is to ground the coaxial shield, "short circuiting" the common mode. Antenna currents flow into such a ground freely, in principle not interacting with noise currents. The best ground for such a purpose will be a earth ground near the antenna and far from utility lines.

Still another way is to block common mode waves by burying the cable. Soil is a very effective absorber of RF energy at close range.

Unfortunately, none of these methods is generally adequate by itself in the toughest cases. Baluns are not perfectly effective at blocking common mode currents. Even the best balun can be partially defeated if there's any other unsymmetrical coupling between the antenna and feedline. Such coupling can occur if the feedline doesn't come away from the antenna at a right angle. Grounds are not perfect either. Cable burial generally lets some energy leak through. A combination of methods is usually required, both encouraging the common mode currents to take harmless paths (grounding) and blocking them from the harmful paths (baluns and/or burial).

The required isolation to reach the true reception potential of the site can be large. According to the measurements I quoted above, for my site the antenna noise floor is 18 dB below the conducted noise level at 10 MHz. 18 dB of isolation would thus make the levels equal, but we want to do better than that: we want the pickup of common mode EMI to be insignificant, at least 5 dB down from the antenna's floor. In my location the situation gets worse at higher frequencies as the natural noise level drops and therefore I become more sensitive: even 30 dB of isolation isn't enough to completely silence the common mode noise (but 36 dB *is* enough, except at my computer's CPU clock frequency of 25 MHz).

Getting rid of the conducted noise can make a huge difference in the number and kinds of stations you can pick up: the 18 dB difference between the conducted and natural noise levels in the case above corresponds to the power difference between a 300 kW major world broadcaster and a modest 5 kW regional station.

The method I use is to ground the cable shield at two ground stakes and bury the cable in between. The scheme of alternating blocking methods with grounds will generally be the most effective. The ground stake near the house provides a place for the common mode noise current to go, far from the antenna where it cannot couple significantly. The ground stake at the base of my inverted-L antenna provides a place for the antenna current to flow, at a true ground potential relative to the antenna potential. The buried coax between these two points blocks noise currents.

> There has been some discussion of grounding problems on this and related echos. I believe it has been mentioned that electrical codes require that all grounds be tied together with heavy guage wire.

I'm no expert on electrical codes, and codes differ in different countries. However, I believe that any such requirement must refer only to grounds used for safety in an electric power distribution system: I do not believe this applies to RF grounds.

Remember that proper grounding practice for electrical wiring has very little to do with RF grounding. The purpose of an electrical ground is to be at a safe potential (a few volts) relative to non-electrical grounded objects like plumbing. At an operating frequency of 50/60 Hz, it needs to have a low enough impedance (a fraction of an ohm) that in case of a short circuit a fuse or breaker will blow immediately.

At RF such low impedances are essentially impossible: even a few centimeters of thick wire is likely to exhibit an inductive impedance in the ohm range at 10 MHz (depends sensitively on the locations and connections of nearby conductors). Actual ground connections to real soil may exhibit resistive impedances in the tens of ohms. Despite this, a quiet RF ground needs to be within a fraction of a microvolt of the potential of the surrounding soil. This is difficult, and that's why a single ground is often not enough.

> A little experimentation with my radio showed that the chassis was directly connected to the third (grounding) prong of the wall plug. I am concerned that by connecting my receiver to an outside ground I am creating a ground loop that involves my house wiring. Can you comment on this?

Yes, you have a "ground loop". It's harmless. In case of a nearby lightning strike it may actually save your receiver. My R8 isn't grounded like that, so I had to take steps to prevent the coax ground potential from getting wildly out of kilter with the line potential and arcing through the power supply. I'm using a surge supressor designed to protect video equipment: it has both AC outlets and feedthroughs with varistor or gas tube clamps to keep the various relative voltages in check. Of course the best lightning protection is to disconnect the receiver, but I'm a bit absent minded so I need a backup.

> This may seem like a trivial point but I recently discovered that the main ground from the electrical service panel in my house was attached to a water pipe which had been painted over. I stripped the paint from the pipe and re-attached the grounding clamp and I noticed a reduction in noise from my receiver.

Not trivial. Not only did you improve reception, but your wiring is safer for having a good ground.

I suspect part of the reason I see so much noise from neighbors' appliances on my electric lines may be that my house's main ground wire is quite long. The electrical service comes in at the south corner of the house (which is where the breaker box is), while the water (to which the ground wire is clamped) enters at the east corner. All perfectly up to code and okay at 60 Hz, but lousy at RF: if it was shorter, presumably more of the noise current would want to go that way, and stay away from my receiver.

> I am also a little confused by what constitues an adequate ground. I have read that a conducting stake driven into the ground will divert lightning and provides for electrical safety but that RF grounding systems have to be a lot more complex with multiple radials with lengths related to the frequencies of interest. Is this true?

Depends on what you're doing. If you're trying to get maximum signal transfer with a short loaded (resonant) vertical antenna with a radiation resistance of, say, 10 ohms, 20 ohms of ground resistance is going to be a big deal. If you're transmitting 50 kW, your ground resistance had better be *really* tiny or things are going to smoke, melt or arc.

On the other hand, a ground with a resistance of 20 ohms is going to be fairly effective at grounding a cable with a common mode characteristic impedance of a few hundred ohms (the characteristic impedance printed on the cable is for the differential mode; the common mode characteristic impedance depends somewhat on the distance of the cable from other conductors, but is usually in the range of hundreds of ohms). Of course, if it was lower a single ground might do the whole job (but watch out for mutual inductance coupling separate conductors as they approach your single ground).

In addition, a ground with a resistance of 20 ohms is fine for an unbalanced antenna fed with a high

impedance transformer to supress resonance. Such a nonresonant antenna isn't particularly efficient, but high efficiency is not required for good reception at HF and below (not true for VHF and especially microwave frequencies).

Much antenna lore comes from folks with transmitters who, armed with the "reciprocity" principle, assume that reception is the same problem. The reciprocity principle says that an antenna's transmission and reception properties are closely related: it's good physics, but it ignores the fact that the virtues required of a transmitting and receiving antenna are somewhat different. Inefficiency in a transmitting antenna has a direct, proportional effect on the received signal to noise ratio. On the other hand, moderate inefficiency in an HF receiving antenna usually has a negligible effect on the final result. A few picowatts of excess noise on a transmitting antenna has no effect on its function, but is a big deal if you're receiving (of course, one might not want to have transmitter power going out via unintended paths like utility lines: this is indeed the "reciprocal" of the conducted noise problem, and has similar solutions).

Appendix: Absolute RF measurements with an R8.

Although the Drake R8's signal strength meter is marked with silly "S" units, the alignment procedure in the service manual actually sets up the meter to an absolute standard, at least sort of. A 60% modulated signal with a carrier level of -73 dBm (which is really closer to -72 dBm in total power including sidebands) is S9. One S unit is 5 dB. This is with 6 kHz bandwidth and with neither the RF preamp or attenuator engaged. I assume this is what they do at the factory.

Now, I don't really know how accurately this calibration is performed, and it certainly can't be more accurate than the flatness of the input passband filters (spec'd at <2 dB p-p). There are also problems because the measurement is actually being made by a peak-responding AGC system rather than an RMS meter. Based on experience with other peak sensing systems, I estimate that the meter probably reads noise power too high by about 3 dB, relative to the carrier power in the test waveform. Therefore, for noise, S9 is about -76 dBm.

On my R8, the linearity of the S-meter calibration is poor at the very low end: S1 is much less than 10 dB below S3. Therefore, for measurements below S3 I do relative measurements and refer them to stronger signals. I have on my NeXT computer an old demo application that gives the RMS amplitude of a signal on the audio input jack. With the R8's AGC turned off and the RF gain set low enough to insure good linearity, this may be used to make quite accurate relative power measurements. You could, of course, use an ordinary AC voltmeter to do this if you have one sensitive enough to read the level of the Drake's audio output (I don't have one).

Considering all of the uncertainties, the numbers hold together remarkably well, better than the likely accuracies in this case (just dumb luck).

For the measurements quoted in my previous message, the receiver's noise floor is -119 dBm. Drake's specs imply that for a 6 kHz bandwidth the noise floor should be below -118 dBm with the preamp off.

According to "Reference Data for Radio Engineers" (Sams, 1975), the wintertime level of natural noise in my area at 10 MHz should be about 32 dB above the thermal reference level: this would produce a noise floor of -104 dBm in this bandwidth with a perfectly efficient antenna. A calculation for a 17 m vertical antenna feeding a high impedance transformer predicts a loss due to mismatch/lack of resonance of 4.5 dB at 10 MHz. My antenna is not a vertical but an inverted L which I presume is slightly less efficient (difficult to calculate). There are also presumably some modest losses in the transformer, the grounds, the cables and the connectors. I wouldn't be surprised if these added up to 3 dB or so. With a total antenna system inefficiency of 7 dB, I'd therefore expect to see an antenna noise floor of -111 dBm, which is, in fact, just what I measure.

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Grounding Systems for Amateur Radio Stations

By: John Wendt WA6BFH

There is quite a bit of talk these days between Amateur **(Ham)** Radio operators about the desirability for **"good grounding systems"** for their home radio stations. I can tell you honestly that this was not so fervently discussed on the radio in earlier decades. My suspicion for this is that there was a better understanding in earlier times, before the advent of our rapid increase in new operators, of the conditions under which a **"grounding system"** would do some good. Lets look at this carefully and critically so that we can assess the desirability of a grounding system for our radio stations.

Safety Grounds vs. Radio Frequency (RF) Grounds

First of all realize that there are two discreet types of ground systems and reasons why a Ham might desire to provide a **"station ground"**. These are for the most part mutually exclusive! In other words one does one job for which it is specifically designed, and one does a different job. You must design your grounding system with this in mind, or one function may inhibit or nullify the other!

One of these grounding systems is what I will generally describe as a **"safety ground"**. This safety ground is installed to reduce the risk of electrocution or radio equipment damage by short circuited **"power mains"**, or from lightening strikes to the antenna or **"feedline"** system. A safety grounding system is certainly a good Ham Radio **"engineering practice",** although it is usually considered as of secondary importance to an **"RF ground"!** In portions of our country where dramatic lightening storms are common this preference is probably reversed. I can tell you that even in the portion of Southern California that I live in, lightening strikes from annual spring storms, or even the thought of accidental **"short circuit"** electrocution are enough of a concern that I have a safety ground system for my station. I have designed it though to operate in conjunction with the RF grounding system I use! Lets talk about the considerations of properly designing an RF grounding system. Through the course of this discussion I

will explain how an RF ground can be utilized as a safety ground, and also that a safety ground **should never** be used as an RF ground!

The RF or Radio Frequency Grounding System

A popular misconception is that all radio antenna systems should be grounded to enhance their radio performance. If grounding these antennas was for the purpose of safety, as I have clarified in the earlier paragraphs, this might be true. To think though that grounding an antenna will automatically make it a better communications device is completely wrong! To appreciate this we must think in terms of **"radio frequency wavelength"**, or fractions of wavelength at 1/4 wavelength increments.

On certain lower frequency longer wavelength bands an RF ground is not only a reasonable consideration but, it is fundamental to getting the best communications performance from your radio station. At other higher frequency and shorter **"wavelength bands"** however, RF grounds are either superfluous, or even harmful to overall communications performance! Lets look at this carefully as illustrated in typical and practical scenarios. Keep in mind the concept of thinking about the physical length of the **"grounding conductor"** in relationship to its comparative **"wavelength dimension"!**

The Practical Station Ground

Living in a typical single story ranch style home my choices and availability for a good RF ground installation are relatively simple. The first **"ground rod"** of my grounding system is within 4 feet 2 inches or 1.27 meters of my station equipment. This 50 inch dimension is the entire or total length of the conductor to this first grounding rod. From this first 10 foot deep ground rod, a # 6 AWG. (American Wire Gauge) bare copper wire is buried and runs under ground in a one foot deep trench to a second 10 foot long rod about 15 feet distant. This second rod is then connected to a very elaborately designed grounding system at the base of my antenna tower's foundation. The length of this conductor between the second ground rod and the tower base is 16 feet. The entirety of these ground rods, and the interconnecting conductors, are buried at about 12 inches deep. Take a look at Figure 1 to see this system drawn to scale.

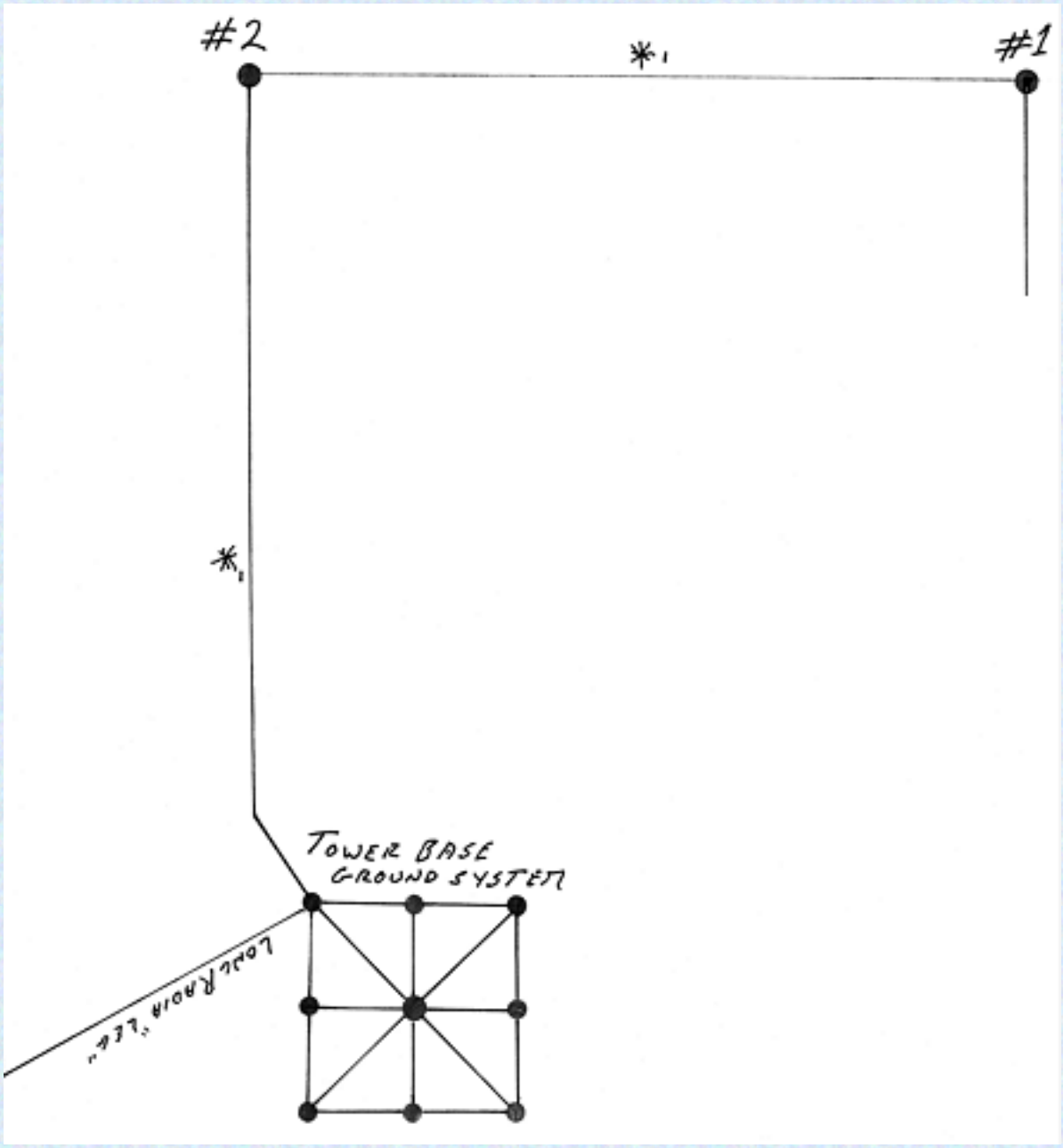

Figure 1.

When you look over this drawing you might think, gee if he says its important to keep the ground conductor as short as possible, why does he have almost 5 meters (15.5 feet) of separation between ground rods. He could have even run a wire from ground rod number 1, straight to the tower base! Well believe me, if I could have, I would have! The combined 15 and 16 foot dimensions surround the perimeter of a building, so that I was forced to take this longer circuitous route! This really though isn't such a bad happenstance!

Having the ground rods this far apart is actually a desirable feature. What is being achieved by this system is to make a **"large surface contact area"** to earth ground. It also functions to reduce **"ground loops"** as well as is possible. Even though I can't prevent all ground loop conditions, I can keep them manageable! I had thought that in my original design that I might also place 10 foot deep rods at the points where asterisks are shown. After testing the system on the MF (Medium Frequency) and HF (High

Frequency) wave-length bands, I found these additional rods to be unnecessary.

Ground Loops (and the worst case scenario)

Ground loops are the major concern and worry to be dealt with as you lay out an RF ground system! Lets lay out a worst possible case scenario of what would be a terrible way to ground a 20 meter (and all harmonically related bands) home station.

Lets say that our 20 meter Ham has his "Ham Shack" in approximately the middle of his house, up on the second story of the house. He uses a nice low resistance and low inductance flat braided strap (lets let him do at least one thing right), which runs from his 1500 Watt amplifier to a 2 foot deep ground rod in his front yard. This braided conductor runs between the floor of the second story and the ceiling of the lower story. It then runs down the outside of his stucco covered home to the ground rod. Its total length is 33 feet long from the amplifier to the ground rod.

Fifty feet (15.24 meters) of coaxial cable goes up between the walls of the second story, is draped over to the top of his tower, and terminates at his TH6DXX beam on this 65 foot tall tower. Lets break down these dimensions in terms of wavelength, and even look at them as they may appear to form a **"current loop"**!

Our rueful Ham friend is using a ground conductor that is almost exactly one half wavelength long and is raised in the air by the buildings structure. Even if this length were reduced to about a quarter wavelength, it would still very nicely couple energy into other wiring in the house, and also the wire mesh beneath the stucco walls. This other wiring could be telephone wires, television cabling, and the 110 volt power wiring in the house. His coax cables shield is a nice **"odd order harmonic"** radiator, as it is 3/4 wavelength long! The ungrounded tower is about a half wavelength tall.

This last point about the tower must be viewed carefully. He might want to ground it as an RF consideration, and he might not! Think about it this way. One ground rod is in the front yard at the end of a half wavelength wire. The shield of the coax cable ultimately goes to ground in the back yard at the tower base at an electrical length of about 7 quarter wavelengths (50 + 65 = 115 feet). This makes another sort of odd order harmonic radiator! If you add into this loop circuit the length of the front yard ground rod conductor, the loop circumference becomes 9 quarter wavelengths. Yet another odd order radiator dimension!

My assessment of this Ham's station would inform me that he is throwing away some of the advantage of his beam and the height of his tower because, all of the radiating conductors strung around the house raise the **"angle of radiation"** considerably higher than it could otherwise be! He is also enhancing the prospects of **"audio rectification"** to his own and to neighbors telephones. He is also enhancing the possibility of **"TVI"** to his television, and maybe to the neighbors as well if they all use a cable TV system.

The Absolute Worst Thing To Do

Maybe though this isn't the worst case scenario? The worst case would be the Ham that tries to use the 3rd wire ground within the 110 volt AC power system that runs though out his house. Then he could really couple RF energy around his house and the neighborhood! **Never ever use this 3rd. wire ground as an RF ground!**

How Should our Ham have done it

1) His station should be on the ground floor of the house.

2) His low inductance low resistance grounding conductor should have considerably less than a quarter wavelength of total length between his amplifier and the first ground rod.

3) If the soil or other conditions in his yard allowed only short length rods, he should have used several of them all tied in a line, and ultimately connected to a **"grounding cage"** or rods near the **outside** of the tower's concrete base.

Important Tip: Don't install the ground cage within the concrete! One reason is that it is connected to the earth via the high resistance of the concrete. The other reason is that a lightening strike to the tower may well blow the concrete block to bits!

When Should an RF Ground Be Used and When is it a Bad Thing

With the advent of the many new MF/HF/VHF transceivers on the market these days, Hams are flocking to frequency bands on which they have little or no experience. Additionally because of the licensing structure, one of the most popular Ham bands in use is the 10 Meter band. I'm sure also that within the coming years, 6 Meters will become even more popular. On 6 meters at least, RFI is very simply controlled.

On one of these two mentioned bands, and also on most bands above 20 Meters, television interference from harmonic radiation becomes a concern that grounding will not fix. In fact most practical possibilities for grounding systems at a typical residential home, will enhance this TVI (Television Interference) problem!

Think about it this way, the 2nd. Harmonic of the 28 MHz. band falls right at TV channel 2 (2 X 28 = 56 MHz.)! This is the fourth harmonic of the 20 Meter band (4 X 14 = 56 MHz.)! From this description and the other worst case scenarios that I outlined, you may be thinking, so is he saying not to bother with installing a ground system? Well, that's not at all what I'm saying! I am telling you though that you must appreciate and design your grounding system so that it cannot possibly contribute to either **"fundamental frequency"** radiation, and also not contribute to the re-radiation of harmonics of these upper HF bands.

The prime methods for accomplishing this are to (1) keep the ground conductor shorter than a quarter wavelength on 10 Meters, which is of course the highest frequency HF band. Next we must, (2) make this conductor be at the lowest AC resistance (impedance) possible! We do this by using large surface area (fat) conductors, and **SOLDER** all connections. After soldering, seal them from the eroding effects of weather by using electricians tape and silicon sealers. Anti-oxidizing compounds such as "No-Ox", which are available at most electrical suppliers, are a must for any connection of dissimilar metals! A simpler way to deal with this last mentioned concern is to simply use non-alloy baring copper for the entire system. Copper tubing typically used for plumbing meets all of the above requirement perfectly. Don't use steel or iron pipe for ground rods! Always (3) connect multiple ground rods in a line. This line can bend or zigzag but, it must extend from the nearest to the furthest ground rod or screen in a line. The last physical consideration (4) is to cover as much ground surface area as is practical within the constraints of your yard or acreage. A minimum RF ground system will use at least three 8 foot or longer ground rods. If you can't sink a rod that long, than you must use many more shorter rods, or bury a splayed out radial system of wires. In this last case you must terminate this wire radial screen within less than a quarter wavelength at 29.7 MHz. (which is 94.5 inches or 2.4 meters).

Thinking conversely, if a grounding system becomes a touchy prospect at the higher bands, might it be a better consideration and benefit at the lower bands such as 160 meters? If you answered this question absolutely, you win a gold star, and Ham of the year award!

On lower frequency bands, both RF noise level, and also **"ground wave"** signal propagation become an important concern. Just to give you a clue, AM broadcast stations invest huge amounts of money in designing and installing their antenna grounding systems. Of course, we are nearby neighbors, almost kissing cousins you might say of the Broadcast band that ends just below 160 meters! On this band doing everything you can to enhance the signal is preeminent within the stations design. It even becomes easier to install a good grounding system than it is to install good antennas!

Think about it. Even if you could put an antenna that is 260 feet long, 100 feet above the ground, it would still be only .19 wavelengths above the ground. That would be the equivalent of installing a 2 meter antenna at about 15.6 inches above the ground! My 160 meter **"Zep"** is 48 feet up at its highest point, this equates to .18 wavelengths. This actually places the **"high current point"**, which is main working point of the antenna, at 37 feet. Thirty-seven feet is about .14 wavelengths above the ground at 1.8 MHz. Pretty low, little more than 1/10th of a wavelength! By contrast, a good ground system is easy! Good RF ground systems really come into their own, as far as showing worthwhile value, on the bands below 30 meters!

Lets wrap up these concepts

An RF ground can be used as a safety ground but, a safety ground is often the worst sort of RF ground.

Multiple ground rods provide the most earth area covered, and consequently, the best sort of grounding

system.

If soil is so rocky that even hydrologically installed ground rod tubes aren't possible, a screen or web of wires below the ground surface will work well. Always use metalurgically sound, true copper wire. No alloys! Cover the largest surface area possible.

Tip: These wires can be installed in a radial fashion, by making slits in the earth or lawn with a sidewalk gardening edger. You then stuff bare copper wire in the slits. In this sort of installation wire size can be reduced to save cost. Use large enough wire to provide good physical strength, lets say #16 AWG.. Be certain that it is **"soft copper"**. A good test is to heat the wire with a gas flame, a propane cigarette lighter works great. If the wire is truly soft copper non-alloy wire, it will quickly turn green, then black, and loose all of its tensile strength. Another test would be to bury a piece of this wire for a week or more. When you dig it up, if it shows signs of oxidation or pitting, its not pure copper!

A ground cage surrounding a tower foundation provides a good final terminus point for the grounding system. It also provides for the prospect of **"shunt feeding"** the tower as a multi-band vertical antenna.

Why use a Ground Cage

For the Ham that is dedicated enough to our avocation to want the best engineered station possible, an antenna tower of considerable height is a must! The best radios in the world will not out perform poorer radios using an efficient antenna!

If you are going to install a tall tower (at least a half wavelength tall in the center of the HF spectrum) your going to have to dig a deep hole for the towers concrete base. The kind of tower I'm speaking of as a minimum will require at least a 6 foot deep hole that is 3 feet per side. I think its advisable though to install a foundation for a larger tower. You never know, you may someday want a nice 90 or 120 foot tall tower. This sort of tower requires at least a 9 foot deep hole that is 4 or 5 feet square. After such a hole has been dug, putting in a ground cage is simple!

First install the longest feasible ground rod in the center of this hole. Hopefully this rod will be at least 8 feet long. Even if you can't install a rod this long into the center bottom of the hole, a shorter rod still adds its length to the wire cage that will reside in the hole.

The cage itself is fabricated from #8 AWG. copper wire. You may visualize this cage as 8 lengths of wire that extend upward from the central ground rod, and reside at each corner of the hole, and the four sides of the hole. It is best to keep the wire contiguous however. If you were to section it into individual lengths, its net resistance (or AC impedance) would be higher than if it is kept as a continuous length. I hope that my verbal description will reveal an assembly that extends 17 feet into the ground (assuming a 9 foot deep foundation hole, and 8 foot rod) and covers a large surface area around the sides of the hole.

After this wire assembly is in place, install the towers steel work base frame. Next before filling the hole

with concrete, install plastic sheeting over the wire. Now when the hole is filled with concrete, the plastic sheeting will insure that the wire is pressed against the dirt sides of the hole, and not encased by concrete. Two ends of the wire can be left to extend above the top of the hole. These will later be attached to the bottom of the tower.

The rod at the bottom center of the hole also provides a solid anchor, to keep the steel work from "floating" or moving about while the concrete is being poured. To accomplish this use Nylon rope and turnbuckles to tension down the steel frame. Nylon rope is used so that no metal work encased within the concrete is electrically connected to the ground system. High voltage static or lightening will now flow only to dirt earth surrounding the concrete. Raise the steel frame above the dirt bottom of the hole with cement foundation piller blocks. If the ends of the steel frame contact the dirt at any point, they will eventually rust and totally disintegrate!

Let's review the best things to do

1) Always use the shortest and largest surface area conductor feasible. This can be wide metal strap material, or large diameter wire or tubing.

Helpful Tip: I use 1/2 inch diameter copper tubing as my ground conductor between rod #1 and the radios. The **"ground bus"** at the back of the radio bench is yet another piece of tubing, with short braided cables soldered to it for attachment to the radios, amplifiers, and antenna **"Trans-Match"**. The ground rods themselves are also copper tubing which were hydrologically sunken (this is a technical way of saying, hook up a garden hose to the tube and let the water suck it into the ground). This method will push aside even grapefruit size rocks as it burrows into the ground.

Copper tubing is much less expensive per foot than is large diameter copper wire (I wish I had thought of that when I installed my system, because it's easier to solder as well). It has a larger diameter than most wire which lowers both its impedance and inductance. It can be easily soldered together using plumbing fittings, just as you would install water pipes. If you want to make a real bang up job of things use Silver solder to lower the impedance even more! Seal and weather proof all soldered joints!

2) Use several ground rods, and cover a large surface area. A large surface area is more important than ground rod depth! Longer rods are desirable though when possible. If it is possible, place ground rods in an area that is often irrigated. Wet soil improves soil conductivity, and helps reduce radio frequency noise.

3) Think about and sort out possible ground loops on various wavelength bands. Design your system to avoid ground loops on any band on which you operate.

Terms you should become convivial with, and use in your conceptual thinking process

Angle of radiation: The term "angle of radiation" and "signal gain" are virtually the same thing. A good

antenna that is high in the air, and also isolated from other random radiators provides the lowest radiation angle, and consequently, the furthest signal propagation.

Audio rectification: This is the condition that prevails when transistor and diode circuitry within telephones or other audio devices are within the "near field" of a radio signal. This same sort of diode junction detection can even be observed with the false switching of infrared detector outdoor lighting. In this last example, the RF signals forward bias the lamps switching circuitry.

****Current loop:*** Any time a wire or assembly of wires form a loop, a current generator is formed. Whenever you impose generated current across a resistance, you produce voltage. In the case of this articles premise, this voltage represents unwanted signal radiation!

Feedline: This term depicts the generic and proper description of conductors that connect an antenna to the radio station. This "feedline" might be either coaxial cable, balanced "Ladder line", or even single conductor line (as in the feed system for a "Windom" antenna, and also "G-line").

Fundamental frequency: The fundamental or prime frequency at which an oscillator or "Exciter" operates. An example might be an oscillator or Exciter operating at 3.5 MHz. (See Harmonic)

Ham: The name by which radio experimenters identified themselves, prior to the implementation of Amateur Radio pursuant to the Federal Radio Bureaus, "Communications Act of 1934".

Harmonic: The multiple of some fundamental frequency. The second harmonic of

3.5 MHz. is 7.0 Mhz., or 2 X 3.5 MHz. = 7.0 MHz.

High current point: Radio frequency currents reverse every half wavelength. The high current point of maximum signal radiation is at the center of a half wavelength antenna.

Good engineering practice: Amateur Radio operators are required by law to construct and maintain their radio stations to the best "state of the art" good engineering practice.

Good grounding system: A grounding system that provides the lowest possible resistance or Alternating Current (AC) impedance, has the lowest AC inductance possible, and is designed so as to limit the re-radiation of fundamental or harmonic frequency energy.

Ground bus: A common point connection terminus that embodies the design criterion of a good grounding system.

Grounding conductor: This is the connecting link between the radio station equipment bus, and the nearest point of earth grounding connection.

Ground rod: The term used to describe the simplest device associated with any sort of electrical "earthing" practice.

****Ground loops:*** Intrinsically the same thing as a Current Loop.

Grounding system: The entirety of a conceived and engineered plan for providing "earthing" perameters for radio or other electrical requirements.

Ground wave: One electrical component of the physics of low frequency radio propagation. In Amateur Radio practices, the 160 band is the only wavelength band that exhibits significant ground wave signal propagation.

Large surface contact area: In the context of this article, this term implies the best "current conductivity" to earth ground. In Alternating Current (AC) circuits (ergo radio) large surface contact provides both low inductance, and low impedance. (See also "Skin Effect")

Near field: The near field of radio frequency energy is that radiated energy within a several wavelengths of an antenna or other radiating source.

Odd order harmonic: The direct definition relates to the third, fifth, seventh, ninth, eleventh, and thirteenth multiples of a fundamental signal etc. A readily available example of an odd order harmonic in Amateur Radio practice is the means by which a 441 MHz. signal can be generated from a 147 MHz. source (147 X 3 = 441). Another way of saying this is, the 70 centimeter band falls at the 3rd. Harmonic of the 2 meter Amateur Radio wavelength band. Odd order harmonics are easily propagated in contrast to "Even Order" harmonics which will self cancel.

Power mains: The generic description of the primary power distribution source (117 Volt Alternating Current, or 220 VAC etc.)

Radio frequency wavelength: The unit of radio signal measurement that embodies both the legal requirement of Amateur Radio station operation, and also depicts the physics by which radio signals propagate. Radio wavelength is determined by deviding frequency into the velocity by which that signal energy travels. Where V = Velocity or 300,000,000 meters per second, and f = Frequency (of oscillation) of the frequency in use. Example: V f = λ 300 / 50 MHz. = 6 meters

RF ground: A grounding system that by inference implies the qualities of low impedance, and low inductance.

Safety ground: A grounding system that by inference implies only a better electrical path to ground than the person or equipment being protected.

Skin effect: Alternating Currents, such as radio frequencies, flow only on the outside surface of conductors. By inference in this application, larger surface area equates to lower impedance and improved conductivity.

Short circuit: The situation that arises when a person or other conducting object is exposed to an electrical circuit between that circuits source and load. The "short circuiting" object in this case, becomes the interim load.

Shunt feed: In the context of this article, this implies a system for "feeding" radio signal energy into a grounded metal tower, to make that tower function as an antenna.

Soft copper: The term soft copper refers to copper that is not produced with other metal content. In other words, it is just plain copper, not an alloy.

Hint: To expand upon the conceptual idea contained within the use of such metals, think about the concept of using other metals of low resistance! Just as a thought teaser, if the cost was not so exorbitant, would using silver plating on antennas or other RF conductors be desirable?

Station ground: A radio station grounding system where both electrical safety precautions as well as radio frequency transmission optimization have been provided.

Trans-match: This is the proper term for a transmitter to antenna system impedance matching coupler. This is often errantly referred to as an "Antenna Tuner". If the word Antenna *System* Tuner was inserted, the errant quality of the term would be mitigated.

Wavelength bands: A political distinction for allotting and cataloging portions of radio frequency spectrum. This term though embodies the physics by which radio signals propagate. (See: radio frequency wavelength)

Wavelength dimension: This term refers not only to unitary or single units of frequency wavelength such as 10 meters or 2 meters, it more importantly refers to thinking in terms of a dimension length. Example: 10 feet is .28 wavelengths at 28 MHz. (or a bit longer than a 1/4 wavelength).

The formula to calculate this is:

V f =

300 / 28 MHz. = 10.7 meters

10.7 X 39.37 = 421.8 inches

421.8 / 12 = 35.15 feet

10 feet / 35.15 = .28 wavelength

Where V = Velocity of signal propagation, or 300,000,000 meters per second, and 39.37 = 1 meter as measured in inches

This same 10 foot physical length of wire becomes .253 wavelengths (still a bit over a quarter wavelength) at the 12 meter band, .508 wavelengths on the 6 meter band, and 1.46 wavelengths on the 2 meter band.

For our purposes in this article, anytime a piece of wire at your station approaches a 1/4 (.25) wavelength or longer, it isn't a ground conductor anymore! In fact it becomes a **"radiating element"**, or you might say an antenna in its own right. This is bad news for a ground system!

Radiating element: a radiating element is any metallic conductor that is within the **"near field"** of a radio frequency energy source. This is usually specific length component elements of an antenna. It can be any piece of metal that is a significant percentage of the energizing wavelength frequency. Examples of such **unwanted** radiating elements might be, rain gutters, wire guy lines, or grounding conductors, any of which would have to be a quarter wavelength or longer at the energizing frequency.

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Tech Bench Elmers λ *© 1996-98*

K3MT
presents . . .

The *GRASSWIRE*
another approach to hidden HF antennas

April, 1997

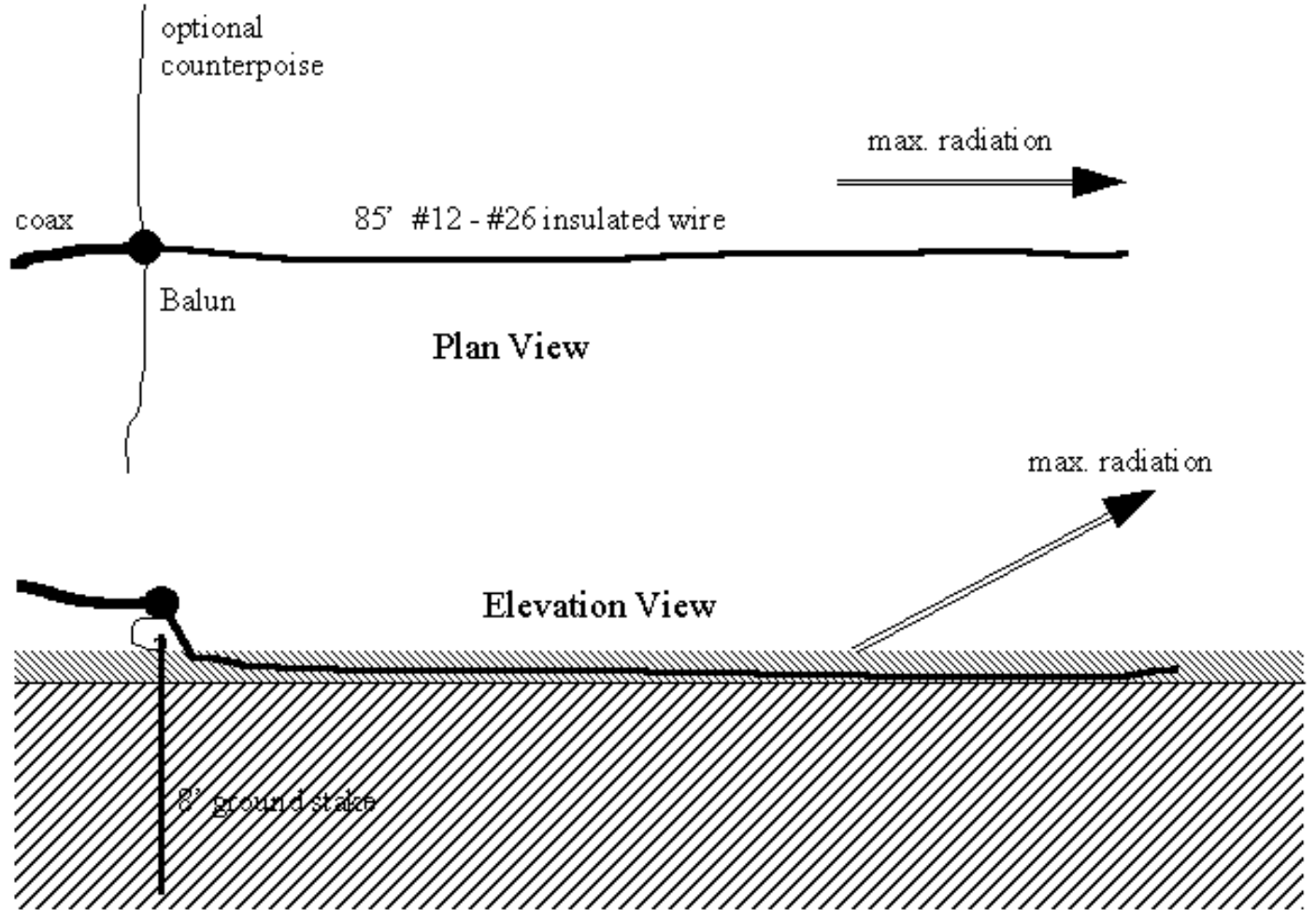
Deed restrictions got you down? Neighbors intimidating your tower plans? Need a really *easy* portable HF antenna? Then the *grasswire* may be the answer! Virtually invisible, lightweight, and compact (you can carry one in your hip pocket), this antenna works! It has been used by K3MT in various installations for more than 10 years.

Read on - and listen to the "experts" telling you that this is hogwash, that an antenna like this can't work. But it does. And *true* experts, who have taken a decade or more to come to grips with the intricasies of Maxwell's Math, know *why* it works.

This antenna will *not* out-perform a yagi, or a decent dipole up a half wavelength. Not in gain or signal strength, at least. But it will survive an ice storm, wind storm, and is practically immune to lightning. And it doesn't need a large tower or tall support. I deploy one from my hip pocket at times - the balun to match it is larger than the antenna!

THE GRASSWIRE - IN BRIEF

What is it? Put simply, it is an end-fed, longwire antenna that is laid right on the grass. Hence the name. The original grasswire used by K3MT in the summer of 1988 was just 204' of #18 AWG magnet wire laid along the property line, anywhere from 1" to 6" above the ground. This sketch shows plan and elevation views of a typical installation. Both an 8' ground rod and optional counterpoise wires are shown. Use one or the other. Both are not needed.



These antennas are largely resistive, with values ranging from 150 to 500 ohms or so on average ground. They have been used successfully on the average soils northwest of Washington, DC, on the sandy soils of the Cape Canaveral, Florida area, in the rocky, shale soils of the mountains in Somerset county, PA, and on river bottomland of Allegheny County, PA. One was used with great success by K3MT/VP9 in Southampton, Bermuda - the object of nightly pileups on 30 m CW for four nights.

REFLECTION AND THE BREWSTER ANGLE

The skeptic in you will doubt that such low antennas can work. After all, its image in the ground radiates and cancels out all radiation. True - if the ground is perfect. But nothing is perfect! The grasswire radiates *vertically polarized* off the *end* of the wire. Extensive monitoring tests with wires laid along the great circle route toward WWV, and perpendicular to that line, demonstrate the end-fire nature of the antenna. So why does it work?

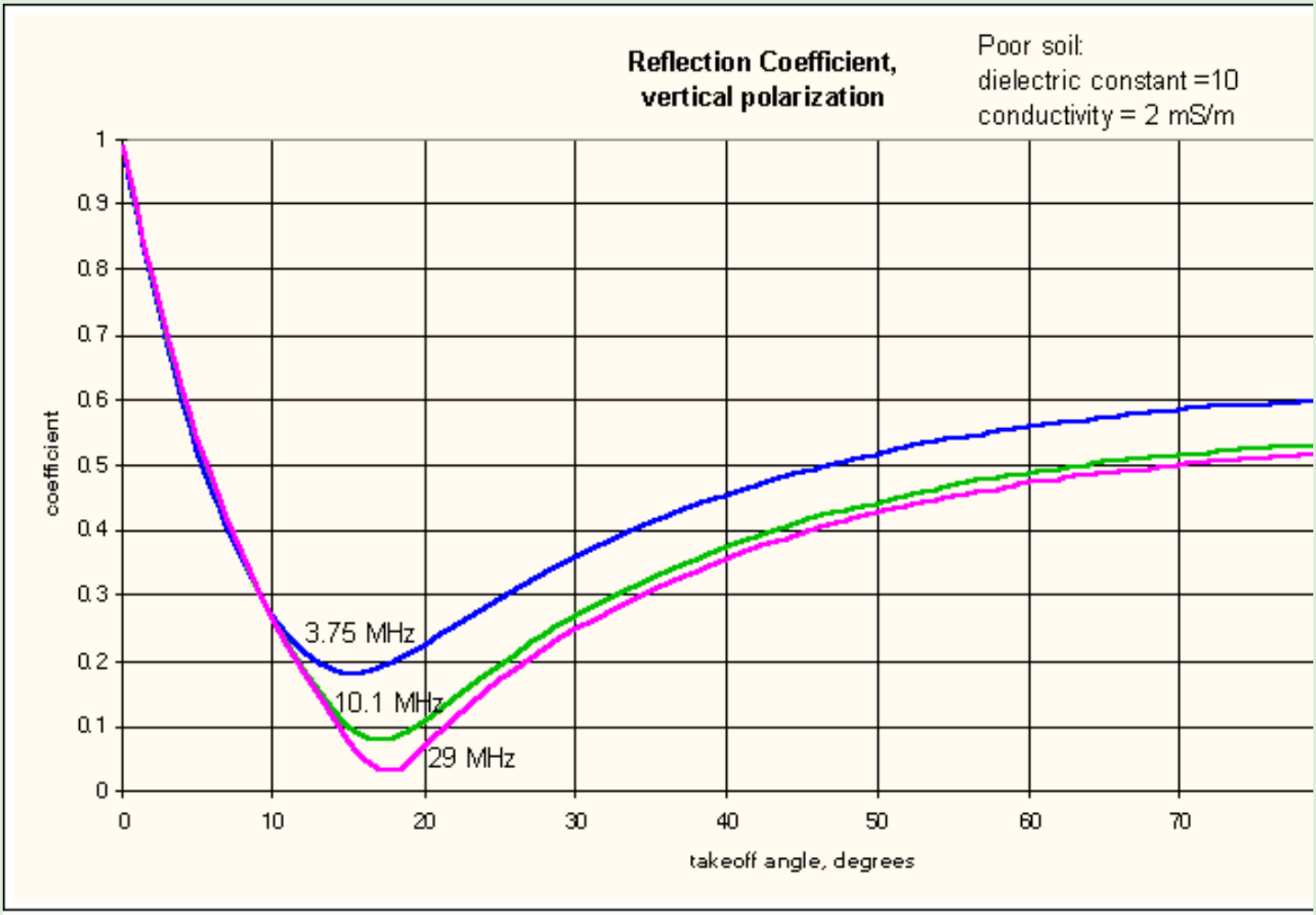
When a plane wave reflects from an air-earth boundary, an incoming ray reflects, giving an outgoing ray. These two, and the line normal to the boundary plane, form

a *plane of incidence.* Solutions of Maxwell's equations differ for the case of the E-field being perpendicular to this plane (i.e., horizontally polarized), and the case when the E-field vector is *in* the plane of incidence. You will probably call the latter "vertical" polarization, although this is technically not correct. Electromagneticists (a.k.a those who practice Electromagical effects) refer to these cases as *normal* incidence (horizontal polarization) and *planar* incidence (vertical polarization.)

For the normal incidence case, reflection is nearly total, with a nearly 180 degree phase reversal. Thus very low antennas neither respond to, nor generate, appreciable amounts of horizontally polarized radiation. But for the planar incidence case, the reflection varies in strength considerably. At some *takeoff angle* (angle between outgoing ray and the ground) the reflection becomes quite weak, and has a 90 degree phase shift. Near this angle, the sum of direct and reflected rays will have a magnitude as if the antenna were in free space! Of course, at other angles, ground reflection largely cancels the direct ray, and the antenna does not radiate well at all.

A *reflection coefficient* is calculated as the ratio of the electric field in the incoming ray to the electric field in the reflected ray. It varies from one (total reflection without loss) to zero (no reflection at all.) It depends on the takeoff angle, frequency, and the soil parameters (dielectric constant and conductivity.) Here are plots of planar incident (vertical polarization) reflection for typical "good" and "poor" soils.

Reflection Coefficient,
vertical polarization
Good soil:
dielectric constant =20
conductivity = 30 mS/m
3.75 MHz
10.1 MHz
29 MHz
coefficient
takeoff angle, degrees
0
0.1
0.2
0.3
0.4
0.5
0.6
0.7
0.8
0.9
1
0
10
20
30
40
50
60
70



Notice that, at 10 to 25 degrees, the ground reflection is very weak. It also is shifted 90 degrees in phase from the incident ray. Therefore, radiation from the grasswire, off the ends will be about the same as if the ground were not present.

But launching a ray at, say, 15 to 20 degrees takeoff angle, in a direction toward Europe, can be useful! That's what a grasswire does. It is lossy in all directions, but least lossy when exciting the ionosphere for a long-haul DX contact. To demonstrate the point, here's an extract of K3MT's log, for October of 1988, (ahh, glory! Yes, the SSN was good then!) using a grasswire:

```
Date     GMT      CALL     his/my RST     FREQ     Power

OCTOBER

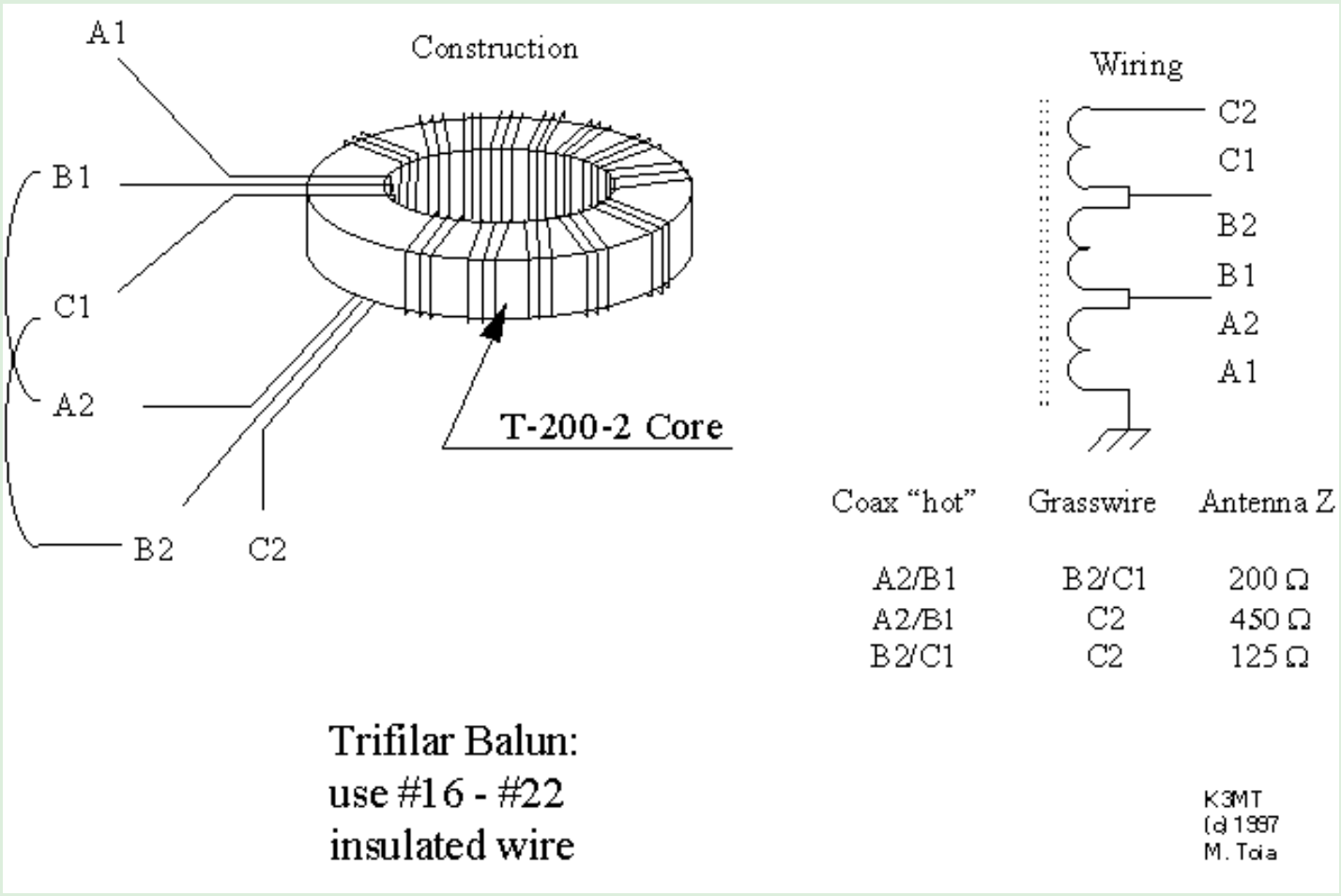
27       1554    SM6DYK     579 / 559    28004      80
         1601    SM0LBR     569 / 439    21007     100      RAY -
STOCKHOLM
```

	2001	W4JBQ	579 / 569	7029	40	JOE - FT WRIGHT, KY
	2141	W8LNJ	579 / 459	28015	80	DAVE - DALLAS, TX
28	0227	W8AO	589 / 569	3547	15	BOB - SILVER LAKE, OH
	1720	G3RFE	579 / 559	21016	100	TOM - BARROW
	1932	G0CBW	569 / 559	14029	50	MEL
	1945	VE2FOU	589 / 559	7032	100	ANDRE - IBERVILLE
	2026	KB7UX	569 / 539	21040	100	RUSS - CHINO VALLEY, AZ
	2100	I2JIN	589 / 559	14022	40	BOB - COMO
	2123	G3JVC	569 / 559	14022	40	JOHN - LONDON
29	2105	WA200JXT	599 / 599	28015	80	ND

Not bad, for a wire on the ground. Notice that contacts were made on 80, 40, 20, 15, and 10 meters. The signal reports are not fantastic. But contacts were made, and ham radio was enjoyed! Five countries were worked in 3 days. And the best part of this setup: *the neighbors never knew that a ham station was on the air!*

FEEDING THE GRASSWIRE

Since this antenna is largely resistive, a simple trifilar balun is all that I have ever had to use. This sketch shows how to make a balun that works:

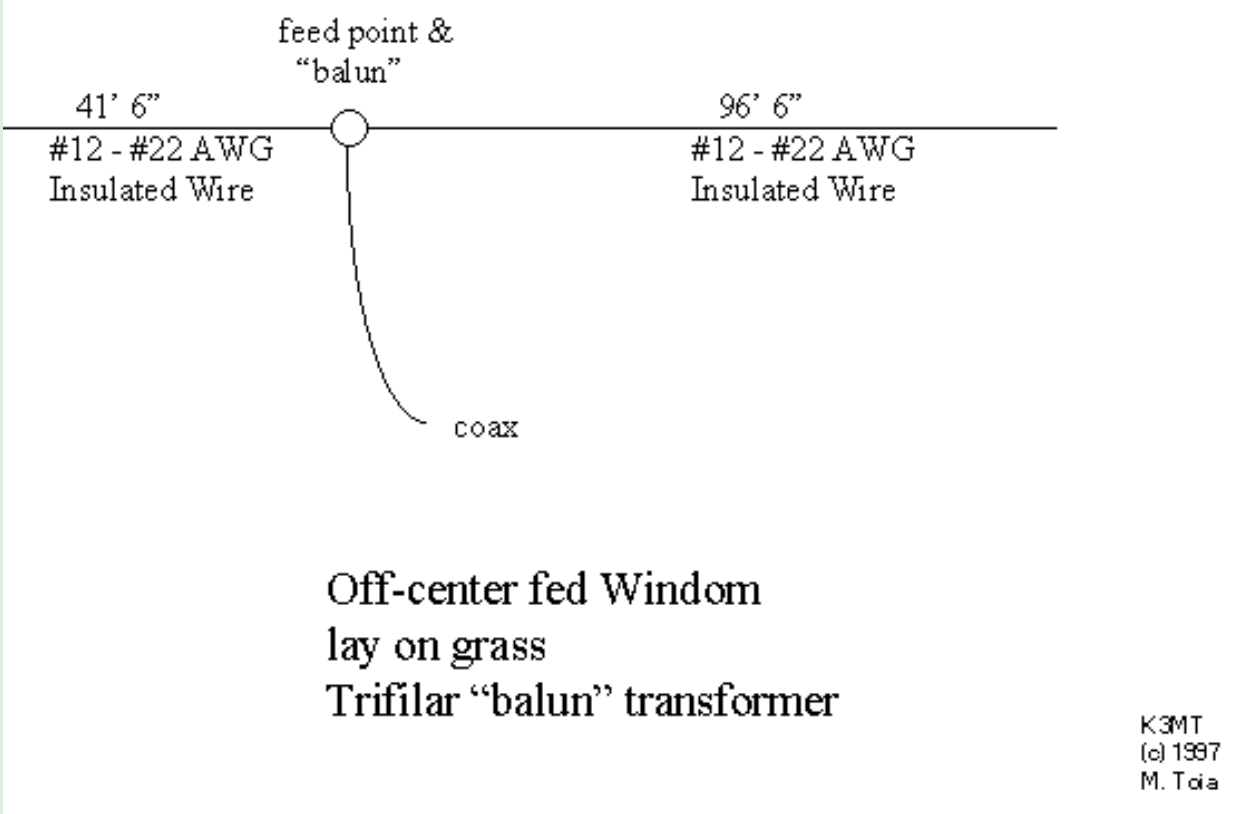


Typically I pull the insulation off of some indoor telephone wiring cable. Four insulated #22 copper wires are inside: discard one of these and use the remaining three. Wind about 16 turns on the core, without allowing the wire to twist (keep the three conductors parallel at all times.)

Notice that this "balun" really matches an unbalanced antenna to an unbalanced transmission line. It is basically a wide-band, three-winding autotransformer. Impedance ratios are as shown on the drawing. Generally it is necessary to connect the coax to either A2/B1 or B2/C1, and the antenna to B2/C1 or to C2. This may change from one band to another, and usually does.

WINDOM IN THE GRASS

I have elsewhere described a windom antenna. While it is usually hung from a pole or in a tree, it works when used in a "grasswire" mode. Just lay it on the ground. Dimensions are repeated here for ready reference.



When I travel, I take one of these made of #22 insulated hookup wire. Since I often set up beside motel parking lots, and often after a day's work, the longer wire is black, and the shorter one is red. This helps me determine which way to point the windom. Remember, though, that it fires *off the long end*. Of course, it fires the other way, too, but usually works best off the long end.

I hope this has given some of you a good case of curiosity. Go out and try one of these ground - mounted wires. They're easy to build. Even the balun is easy to build.

If you must, contact us: we can supply a core, a whole balun, or a whole grasswire windom setup.

For more unusual antennas, visit my web page.

And check out my Book for a dozen topics about HF antennas.
that includes the grasswire and other beverage antennas.

73

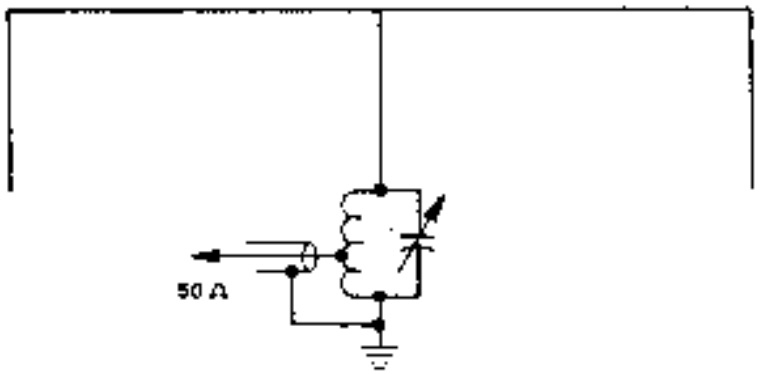
K3MT

Bobtails and Half Squares

These antennas offer big gun DX performance on the low bands, yet they are surprisingly simple to build. Try one- you will like it!

ABOUT THESE ANTENNAS

Bobtails are vertical arrays that look like the letter "E" turned 90° clockwise onto its tails. They have 3 quarter wave vertical elements and are one wavelength long. The Half Square is a shortened Bobtail- it has 2 quarter wave vertical elements and is a half wave long. The tops of the verticals are all connected together by a wire or other conductor. The bottoms of the verticals are all insulated from ground. Bobtails are fed from the bottom of the middle element. Half squares may be fed at either end- it matters very little in the resulting pattern.



These antennas belong to the 'inverted ground plane' family. They work best *voltage-fed,* rather than current-fed. You should not connect coax directly to the tails. Current-fed antennas like the dipole have low impedance feedpoints and are a good match to coax. The impedance at the tips of a dipole is high, however- in the thousands of ohms, like the bobtail. Fortunately it is easy to voltage-feed the bobtail with a simple LC parallel resonant circuit connected between the element tip and ground. With the point of highest RF current a quarter wave up in the air, ground requirements are much less stringent than with ground-driven, current-fed verticals which need many radial wires to work well.

It is customary to use only a small ground screen as the feedpoint counterpoise. A 3 by 5 foot piece of wire cloth is often enough, even on 80 meters. It can be supported a few feet up or it can be laid directly on the ground or even on a roof. It works much better than a ground rod, which may be connecteded to the counterpoise but should be thought of only as a lightning ground. A coax feedline can be connected to a tap near the grounded end of the coil and the tap adjusted for best SWR, or even better, inductive link coupling can be used.

The horizontal pattern of a bobtail is a bidirectional figure 8 broadside to the plane of the antenna. The two lobes are rather narrow- only about 50°. It has a respectable gain of about 5dB over a

single vertical. The half square has a beamwidth of about 60° and a gain of about 4dB. *Apparent* gain for DX (over single verticals or low horizontal antennas) will seem to be greater. Both antennas have good side rejection that helps hear the DX better. The bobtail has a cleaner pattern that better rejects close-in noise and signals (some of your competition in the pileups.) The vertical angle of radiation is low, in the 20° to 30° range, depending on the conductivity of the local earth.

Note that the bobtail is primarily a low bands antenna useful on 30 meters and below. 30 is a transitional band where vertical and horizontal antennas both do well for DXing. For 20 meters and up, almost any horizontal antenna at reasonable height will usually do better than a vertical (unless your location happens to be over salt water.) I did build a 20 meter bobtail once in an emergency when the yagi atop my tower had failed and there was a DXpedition to a rare country coming up. The center tail was voltage-fed, with the feedpoint on my roof at about 17 feet. The stations I worked reported it had a potent signal. There have been favorable reports in the literature as well. Perhaps these antennas are underrated and underappreciated on the high bands. I doubt it is worth bothering with one for 20 meters and above, however, if you are able to put up a tower and yagi.

New bobtail owners often report great success with their antenna. It is designed for DXing, not for close-in work where low, horizontally polarized antennas are more appropriate. Mounted at heights of about 25-30 feet for 30 meters and 35-40 feet for 40 meters, these simple antennas deliver DX performance that rivals horizontal antennas in the 100+ foot class. Furthermore, at these heights it is practical to use trees or simple wood or metal masts to support the wires. Metal tubing or pipe masts can also double as the vertical elements. If you have tall trees as I do, 80 meters is not out of the question. Mounted at 70 feet, the tails of a wire 80m half square will be a few feet off the ground. I have used both a 40 meter bobtail and 80 meter half squares and was very pleased with them. I would not hesitate to build some again, under the right circumstances. They are a hard-to-beat combination of DX punch and simplicity.

More about Bobtails and Half Squares:

Bobtail/Half Square Grounding and Feeding
Beverage and Half Square Bibliography (includes articles by the inventor, W6BCX)
Detailed plans for building matching networks for your favorite bands
Construction notes and important safety issues
User Comments from Bobtail and Half Square owners
Half Square/ Bobtail Articles Critique

Bobtail and Half Square users: Please email your experiences with these antennas to be included in the User Comments section.

EMAIL Send Comments to K3KY

The Low Bands PREV

HOME K3KY's DX Toolbar

Grounding and Feeding NEXT

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Elevated Half Square

KQ6RH

(C) 1998, 1999, 2000

Ray Jurgens

(Up-Dated 2/25/2000)

Elevated Half Square

The elevated half square antenna is a very simple antenna that can be constructed from common materials yet gives significant improvement over a dipole at the same elevation. Ground mounted versions of this antenna are commonly used in amateur radio for 75/80 and 160 meter bands. The elevated version described here provides even greater performance. The Half Square antenna is bi-directional, has a low angles of radiation, is self supporting, is easily erected, and can be rotated with a TV antenna rotator. The Elevated Half Square antenna achieves its high level of performance by not radiating power at high angles of elevation where is leaks through the ionosphere and by having its high current regions near the top of the antenna. Figures 1 and 2 show the radiation patterns in azimuth and elevation.

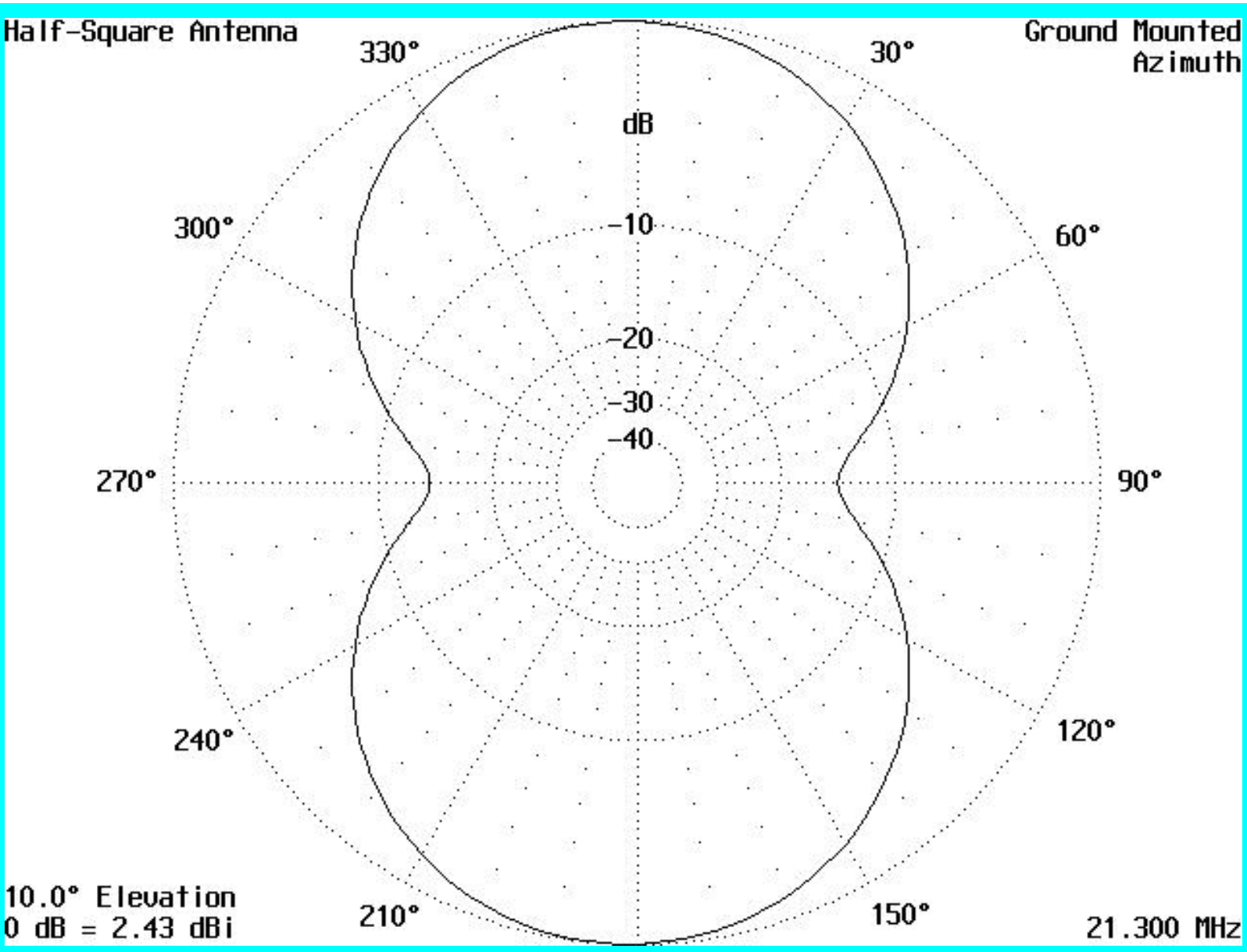

Figure 1

Azimuth Pattern at 10 Degrees Elevation

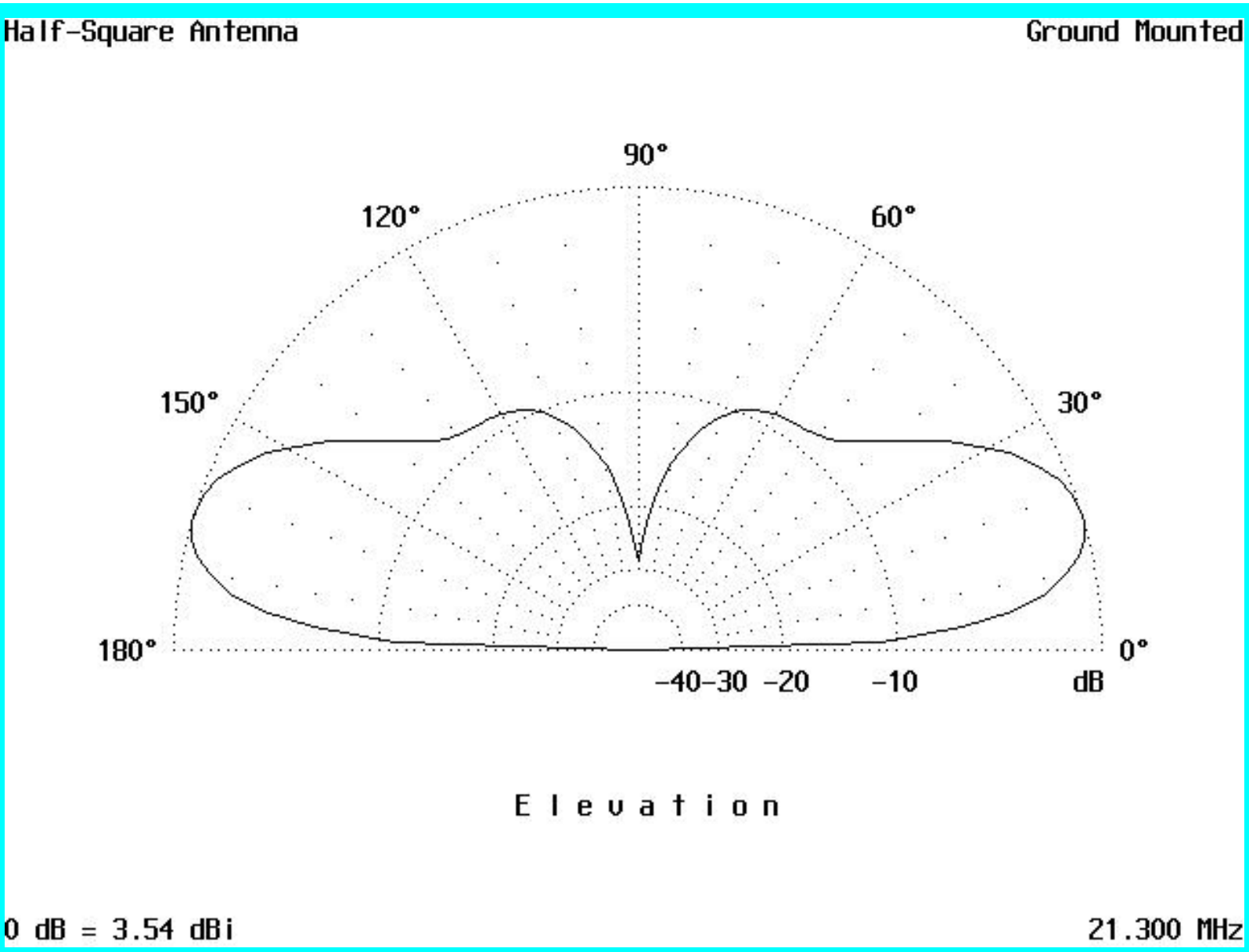

Figure 2

Elevation Pattern for the Elevated Half Square Antenna at 0 Degrees Azimuth

This is one of the best DX antennas you can find unless you just happen to be able to put a 3 element Yagi at 90 to 120 feet in the air. Note that high angle radiation (local QRM) is attenuated by as much as 10 dB, so if you are in the central US, coastal stations will be attenuated, and the Europeans and western Pacific regions will be given 3.54 dBi gain. Forget about using this antenna for local work.

The pictures below show two ways to construct this antenna. There is the neat and light version made with fiberglass structures and the cheap, heavy and ugly PVC version. So take your pick.

Figure 3

Fiberglass Version of the Elevated Half Square Showing Structure and Feed

Figure 4

PVC Version of the Elevated Half Square Showing the Structure

There are a number of tricks to the design of the Elevated Half Square, and our Guide to the Elevated Half Square gives the complete details of the corrections for propagation velocity in dielectric tubes, methods of decoupling the feed line from the radiating structure, the construction of light weight structures to support the antenna and a complete parts list for both fiberglass and PVC structures. These topics are too extensive to cover on a web site, so this guide is listed in our Products Page.

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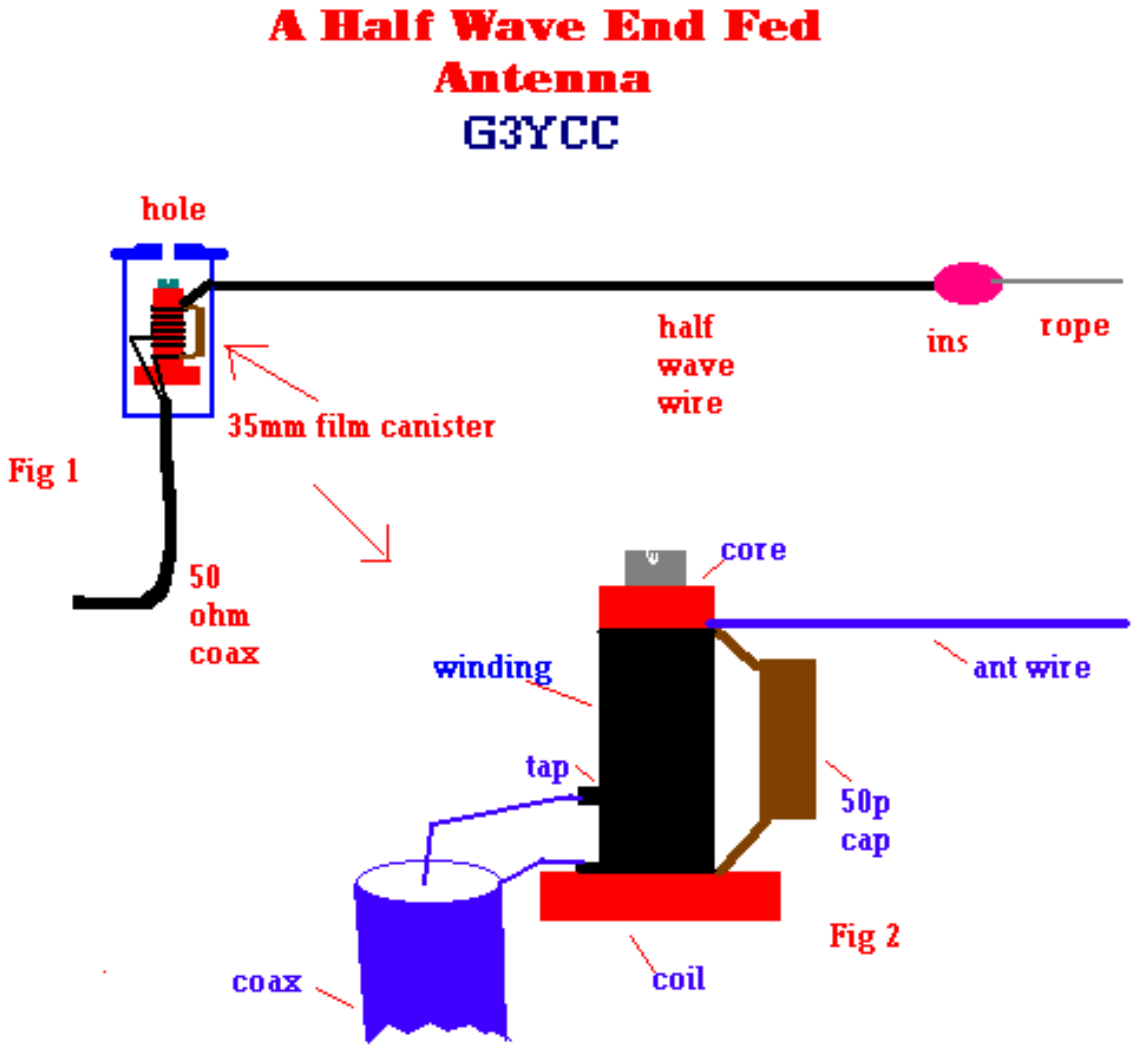
An End Fed Half Wave Antenna

Useful for portable work



This end fed type of antenna was marketted in the UK and is a useful system for the portable set-up. Being a half wave, no radials or counterpose wires are needed.

As the impedance will be high at the end of the wire, some form of matching unit is needed and a simple parallel tuned circuit housed in a plastic film container is fine (figure 1). A hole in the container allows for tuning, by adjusting the ferrite core in the coil for maximum reading on a field strength meter nearby.

The inner of the coax goes to a tap about 1/4 of the way from the earthy end of the winding, to which the braid of the coax is soldered.

As a rough guide, a capacitor of 50pf across a winding of say 30 turns is a starting point for 40m. Some experimenting with turns and tapping point is necessary. The coil former used was about 1/4 inches in diameter as found in old IF transformers.

The antenna wire (1/2 wave long) is soldered to the top of the coil as shown in figure 2.

The drawings illustrate the set up, I hope.

For portable use, throw the wire into a tree or whatever, using thin cord or rope onto the insulator (figure 1), tune for best filed strength reading and off you go!

Try it - it works!

Want to work two bands? Use a slightly bigger container with two tuned circuits which have some sort of terminal each, to which one of their respective wires can be connected.

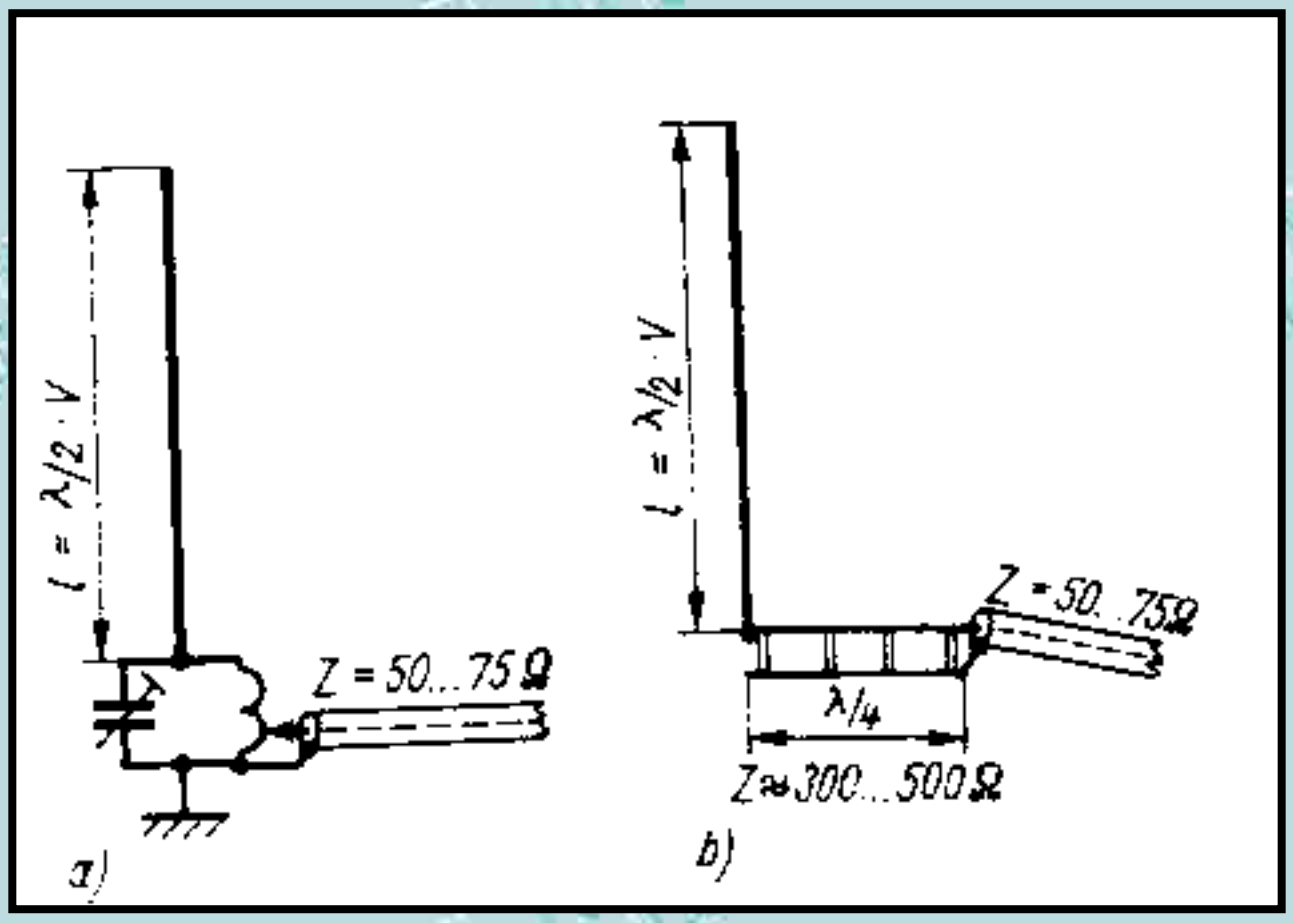
Rough lengths of half wave wires:

80m - 132 feet, 40m - 66 feet, 20m - 33 feet.

Enjoy your portable working and how about telling me what you thought of the above idea - when you have used it?

Back to the first page

End fed-Halfe-wave antenna



This type of antenna has same performance as a dipole, but requires only one single mounting point, which makes this concept ideal for portabel stations. Actual bandwith is more than 2 Mhz with VSWR better than 1:1,7 !
For 14 Mhz you need approx. 10 meters of wire, abt. 4,7m coax-cable 50Ohm or 300Ohm with a connector about 30cm from the shortend end. No tuner is required. We used this antenna on all our holiday - Dxpeditions : 8Q7BZ and in SV8, EA8, SV9, ...

Build this antenna system in less than 30 minutes and have fun and good DX.

top of page | home | propagation |field strength meter | 1270 Mhz antenna | Maldives Dxpedition| Web receivers

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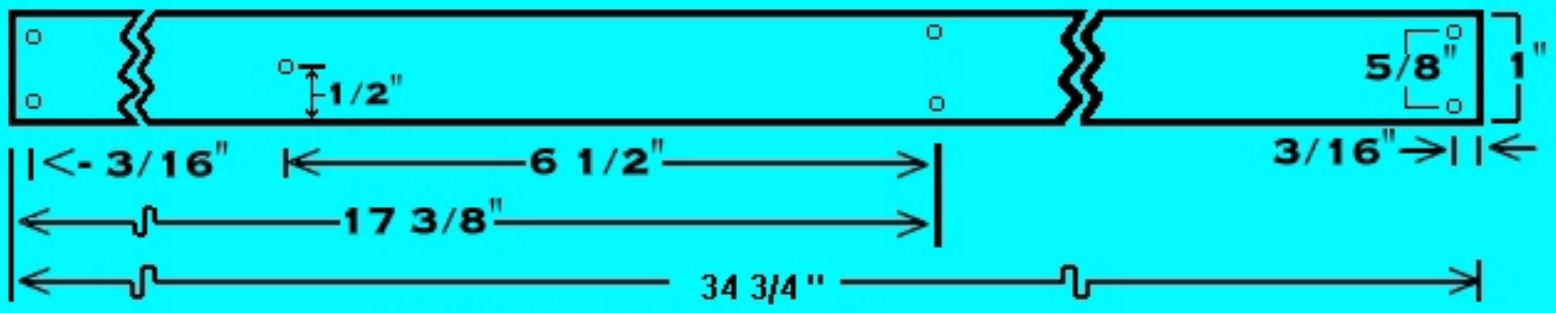
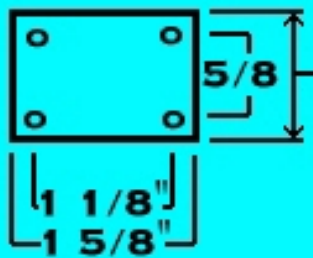
2 METER HALO ANTENNA



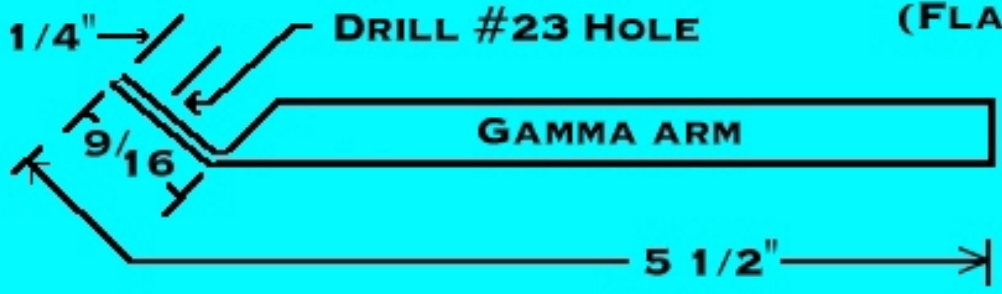
.040 THICK ALUMINUM

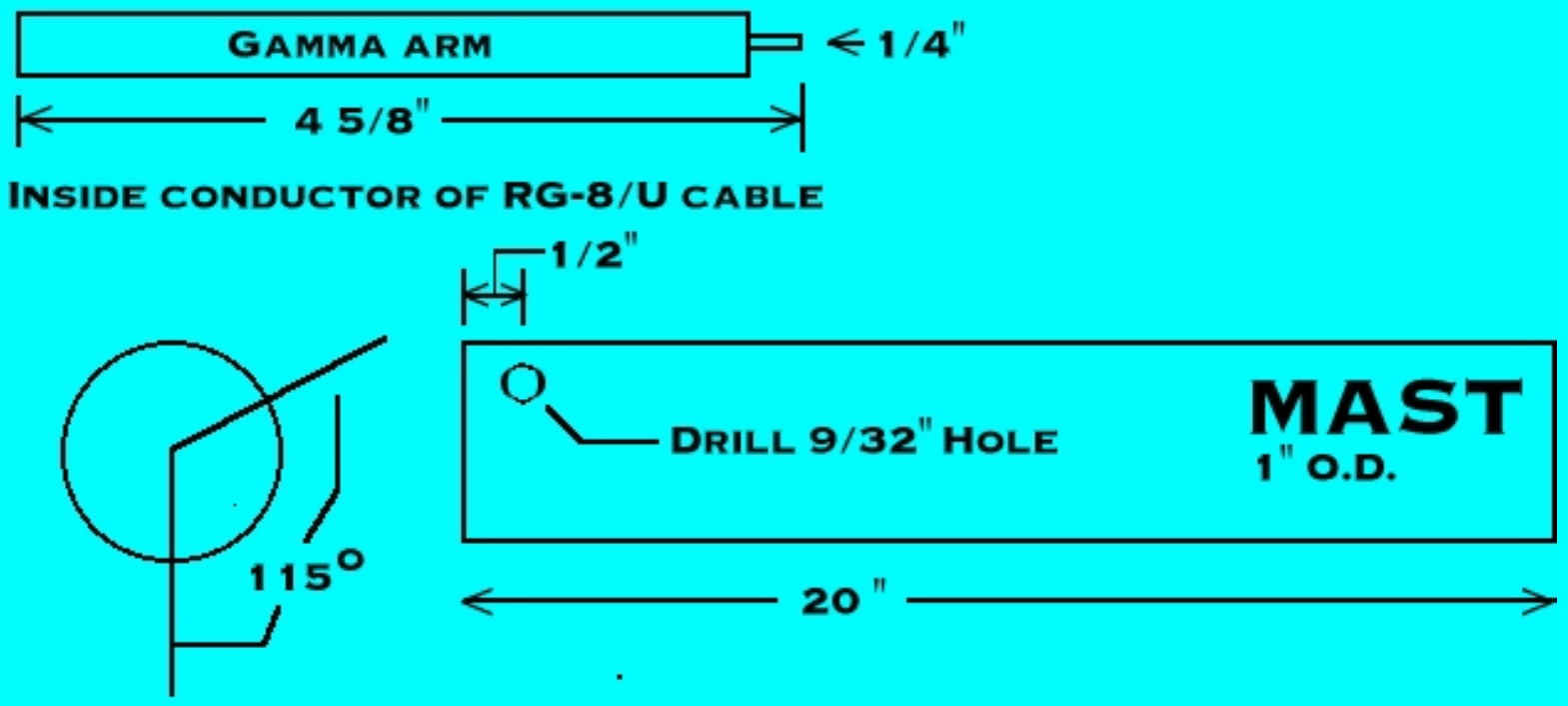
DRILL 7 - #23 HOLES



<------ INSULATOR 1/16" THICK FIBERGLASS

DRILL 4- #23 HOLES





THE HALO ANTENNA IS BUILT OUT OF LIGHTWEIGHT MATERIAL. THE ELEMENT IS MADE FROM FLAT STOCK ALUMINUM THIN ENOUGH SO THAT IT WILL BOW INTO A CIRCLE. THE GAMMA MATCH IS MADE FROM A PIECE OF 3/8 TUBING.THE OTHER PART OF THE GAMMA IS MADE FROM THE INNER CONDUCTOR OF RG-8/U CABLE.IF YOU BUILT THIS ANTENNA AS SHOWN THE SWR WILL BE BELOW 2:1 ACROSS THE 144-148 MHZ.

[Click here to return to front page]

N4UJW

AF4AR/W6ARQ
KC5BBP
"modified"
2 METER SQUARE LOOP

GET ON 2 SSB IN STYLE WITH THIS NEEEEET AND STURDY PLUMBER'S DELIGHT RF PROJECTOR !
LOOK DOWN BELOW FOR ALL THE GOODIES!

BLAST OFF!!
BEWARE OF THE SHOCK WAVE!

ADJUSTING SCREWS FOR TWEAKING SWR
MIKE (AF4AR)
RICK (KC5BBP)
JOHN (W6ARQ)
CAN BE FOUND ON THE 146.67MHZ NASHVILLE,TN AREA MACHINE MUMBLING SOMETHING ABOUT GRID SQUARES AND DX IN THEIR SPARE TIME!

EMAIL THEM
FOR QUESTIONS

Here are the building instructions on the 2 meter square loop.

Credit is given to K0FF for most of this design.

We have added modifications that proved useful in the design.

The mods will be presented under modifications in the design instructions. Special thanks to **W6ARQ** and, **KC5BBP** for thier input on this Antenna.

2 Meter Square Dipole by K0FF

Here is the parts list and dim. sheet for the 2 meter square dipole, made of copper water pipe.

Parts List

1/2 Inch Copper Water Pipe
Long Sides 9 1/2 inches
Open Ends 3 1/4
2 Short Pieces on each side of "T" 4 7/16

4-90 Degree Elbows
2-Copper Caps
1 Copper Tee
Brass Plate for SO239
Gama Tube 4 3/4 of 3/8 Copper Tubing
RG8 is 5 1/2 iches long
Copper or Brass Gama Tube Bracket
SO239

Modification Parts (DO NOT USE STAINLESS STEEL)

6 Brass Nuts-Note Any size
2 Brass Screws at least 1 1/2 long

Instructions The mounting array to be affixed to a standard mobile
mast that presents 3/8 X 24 threads.
Run coaxial cable right to the antenna and connect it to the built in SO-239.
Their are two adjusments on the Antenna to match the coax Imp.
Adjust the Gamma Match Tube for Lowest SWR, then the tuneable stubs move in or out for lower swr.
Do not use Stainless Steel Screws, Soldering them to the end caps is almost impossible.

The 2 meter square loop is a folded Dipole around itself.
The shape is 11" X 11".

Solder the antenna parts together using 90 degree elbows at the corner open ends and mind the gap.
All measurements are critical. The brass plate to hold them form an "L" 1-3/4" tall with a 1/2" lip.
A 5/8" hole is provided 1-1/4 inch from the bend, and attached using stainless steel or brass hardware.
Two small holes are drilled in the lip and mounted to the copper TEE with #6 self taping screws.
Solder the 5 1/2 inch piece of RG8 to the SO-239.
When using the RG8 discard the outer shield and use only the Dielectric. Slip the RG8 inside the 4-3/4 Gamma Tube.
The Gamma bracket should be at least 1 1/2" long brass or copper.
Secure with stainless steel screws right before the 90 degree elbow.
Either side doesn't matter.
The Gap between the Open Ends is 2-3/8".
Adjustments to the gap in or out can be made with the modification
of installing the brass screws on the end caps.

The Antenna is more or less Omni Directional and horizontally
polarized.
It presents a high takeoff angle and is intended for use in the 144.200mhz area.
The Antenna can handle 100 Watts and has proved useful in working some satellites such as AO-27.
The Antenna provides excellant SSB results while mobile and have talked with other stations 200 miles with simularly equipped Antennas.
For further Information on assembly or instruction details Email
the following hams below.
Af4ar6077@aol.com
W6arq@charter.net
Kc5bbp lives too far back in the hills of Tennessee to get daylight...much less email!!!!

I have found this Antenna works very well and, is a solid performer.
Best of luck with your Antenna.

EXPERIMENT! EXPERIMENT! EXPERIMENT!

73

Mike Gunter/Af4ar
Murfreesboro,Tn

See the original design on eham.net

THESE INSTRUCTIONS WILL GET YOU ON YOUR WAY TO 2 METER SSB EXCITEMENT!.

PLEASE DON'T EMAIL HAM UNIVERSE.COM FOR QUESTIONS PERTAINING TO THIS ANTENNA..GO TO THE EXPERTS ABOVE!!

FELL FREE TO COPY ANY OR ALL THE INFO ON THIS PAGE AND LET US KNOW HOW IT PERFORMED!
LISTENING FOR YOU
ON 2 SSB

CREDITS
PHOTOS>>>>>W6ARQ
HARD WORK>>>>AF4AR W6ARQ KC5BBP
PAGE "MASTER">>>>N4UJW
(from an article by KOFF on EHAM.NET)

BACK TO ANTENNA LAB

"TRY IT...YOU'LL LIKE IT!"

DID THIS PAGE LOAD SLOWLY? CHECK THIS OUT

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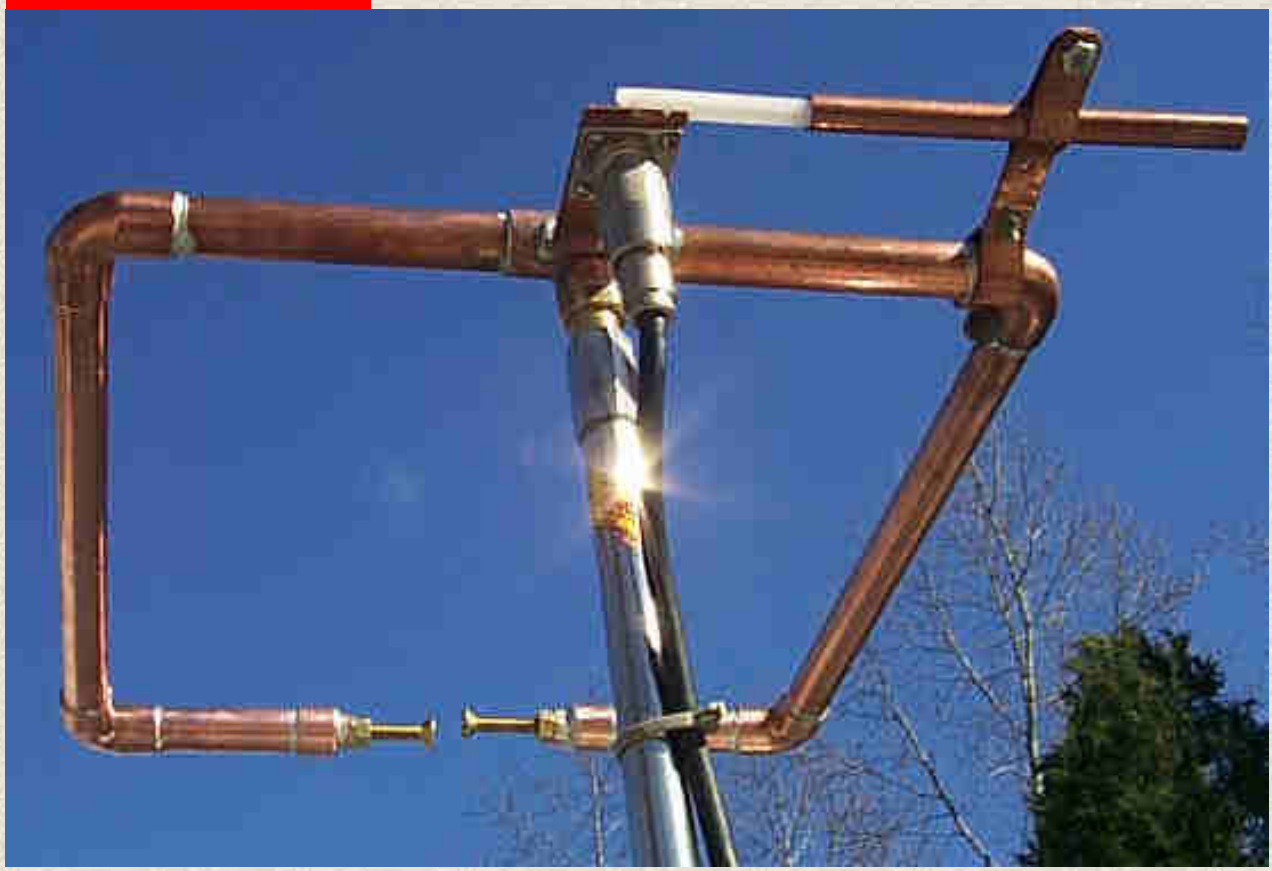
AF4AR USES THIS ANTENNA MOBILE ON AN 18 WHEELER! FILM AT ELEVEN!

AF4AR/W6ARQ
KC5BBP



VIEW FROM THE EARTH

GAMMA MATCH

AF4AR/W6ARQ
KC5BBP

SEE CLOSEUP BELOW >>>>>

AF4AR/W6ARQ
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MOUNTING PARTS

MOUNT ASSEMBLED

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2 METER SSB ANTENNA PROJECT

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OUR THANKS TO ALL! 73 N4UJW**

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HOME BREW 6M HALO

Built for Mobile Use

Jonesy
W3DHJ

The Jones version of the Lawn Chair 6M Halo as viewed thru the rear window of the mobile.
Ain't that just the sexiest thing you've ever seen on the outside of a vehicle?

I read the article: Six Meters from Your Easy Chair by Dick Stroud, W9SR in the January, 2002 QST with some interest. Dick described a 6M Halo antenna that was built from the aluminum frame of a 'generic' lawn chair. It is always a pleasure to peer into the working mind of another homebrewer and scrounger.

However, I put the article and the idea aside, because I had no interest in a 6M *vertical* antenna for this QTH (nor, for mobile operation, either.) And, that was how the antenna was 'portrayed' in the article. However, later in the Summer of 2002, I became Very Interested in improving my meager 6M mobile operation. Heretofore, I had been running 10W out of an ancient IC-551 into a 2M 5/8 wavelength whip -- which 'worked' well as a base-loaded 6M 1/4 wave vertical -- but, probably radiated like crap. I knew deep in my heart that it was not 'fair' to the other op to have to try to copy my peanut whistle.

Harmonic Convergence: It came to pass that one day I was

A: Pondering a better antenna for 6M mobile.

B: Thinking of a better transceiver for 6M mobile.

C: Feeling that the stock market hadn't really hurt me **that**much.

D: Reading the email copy of the "AURORA REPEATER ASSOCIATION SWAPLIST"

There was a used IC-706II listed for sale. Whoa!, I thought. That'll give me 100W on 6M -- it'll do mobile Quite Nicely -- and, to boot, I could dabble in HF mobile. I whipped out my checkbook and became the proud owner of the newest rig in my collection. (I had been using 2 TS-520's, an IC-551, and an IC-251A. I am soooo Last Century.)

While waiting for the UPS Man to drop off my new rig, I had time to ponder just exactly how I was going to 'solve' the 6M mobile antenna problem. I looked at several of the 'commercial' 6M halos that are available. They all looked FB, and the prices were not out of line. But, the Olde Home Brewer and Scrounger was already feeling guilty for **buying** a transceiver. So, I toyed some with the idea of homebrewing the antenna. I played mental games with a few ideas, and eventually circled back to the January, 2002 QST article. Hmmmmmm.... I wondered if I couldn't scratch out something along those lines that would both: radiate horizontally and do 85 MPH?

Well, I think I succeeded. And that explains this website. On the following web pages I hope to explain my thinking, design, material selection, construction techniques, and (yes) mistakes.

These web pages about my homebrew construction of a
Six Meter Mobile Halo are dedicated to the first Jonesy, W3DHJ.
Nearly all my homebrewing and scrounging skills were taught to me by my dad.
I'd like to think he'd be proud of this antenna.

Constructive comments and/or discussions:

6M Halo - Project Beginnings.

W3DHJ Home Page.

HAM UNIVERSE

THE HENTENNA RE-VISITED

The Hentenna was developed by Japanese 6 Meter Hams, JE1DEU / JH1FCZ/ JH1YST in the 1970's and can be designed and built for hf thru uhf and possibly beyond!

After much experimentation, finally, the antenna was developed with good performance, however, it was difficult to explain why the performance was so good, or how it is worked basically at that time. So it was named Hentenna , "Hen" means strange in Japanese.

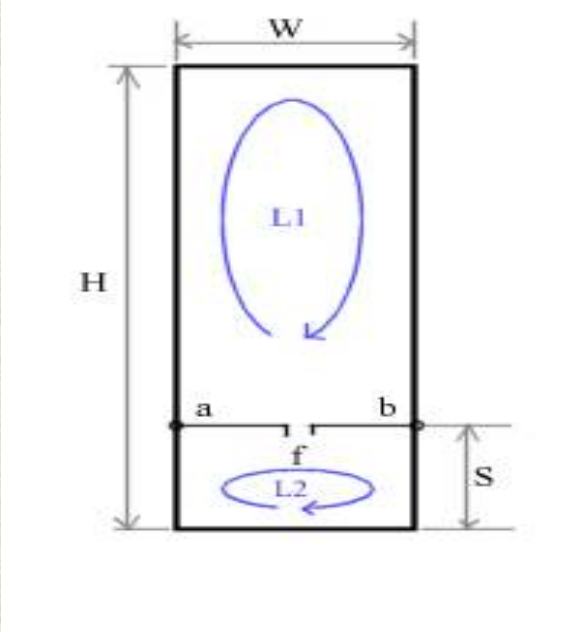
The antenna has good performance and many advantages and it has become very popular in Japan. Many JA stations make it and enjoy it at home or in the field. Some Japanese 6m beacon stations are using the Hentenna antenna.

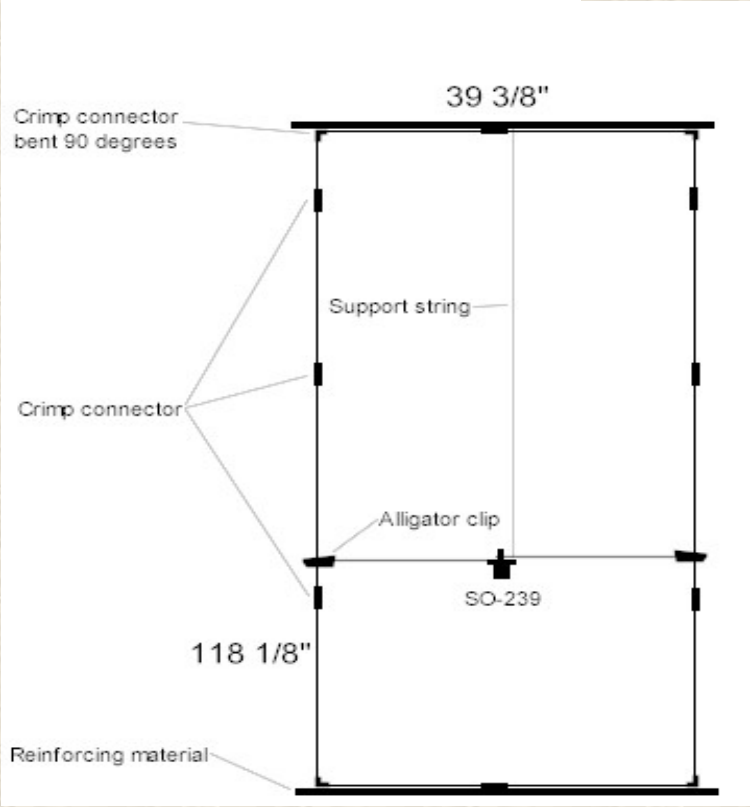
HERE ARE SOME GOOD POINTS FOR THE HENTENNA

1. Good performance
2.5-3 dBd gain
Low angle radiation
*** Total performance is equivalent to 2-3 element Yagi-uda antenna,**
Wide band width
2. Easy to make
It is possible to adjust impedance and SWR perfectly, This means, not so difficult to make!
No special parts are required. You can use any electrical conductor to make the main rectangle.
Broad adjustment ....wide bandwidth
3. Easy to build up
If you use thin aluminum pipe and thin wire, you can make this antenna for 6m very light.
This means, it is easy to put it in a higher position in the air. You can also use light mast for it.
As this is a vertically long antenna, it is easy to install the antenna on a veranda or small space.
It appears to be a vertical antenna but has mainly Horizontal radiation: This is one of the reasons this antenna is "Hen".(STRANGE)

"The following article has been re-edited for the English language from the Japanese site and we discovered some minor errors so corrections were made."

N4UJW





Japan version

English version

Please note in the above English drawings that the Hentenna is basically a loop fed about 1/10 wavelength from the bottom element with 50 or 75 ohm coax attached to the top element of the bottom loop at the center point.

Hentenna Basics

1. Basically 1 1/3 WL Loop antenna around outer edge of antenna
2. L1 works as 1 loop antenna
3. L2 works as matching section
4. Vertical long rectangle has more gain than ordinary square loop and has less impedance. L2 helps the matching and low angle radiation.
5. 3D pattern is like shell of peanut (maximum gain directions looking at you and away from you) so it will be somewhat bi-directional.

How to Adjust

1. Move the "a" and "b" points to adjust swr. (move in equal amounts)
S1 >S Matching Frequency UP (move towards top to increase resonant frequency)
S1 <S (move towards bottom to decrease resonant frequency.

2. SWR may be higher than 1:1.5 at first so move matching points "a" and "b" in small increments up or down the loop until lowest swr is obtained and secure at these points with whatever method you choose depending on your construction materials.

Putting it in the air and on the air

The construction materials you use for the loop will determine how the antenna is supported. It will weigh more if made from aluminum or copper tube and will require a non conductive support mast or structure to attach it even if made from wire. Nylon cord or rope or other non conductive material can be used for support at the four corners.
A length of pvc pipe can be used as support for the top, bottom and coax feed point elements with the side wires strung between them or can be used to completly enclose the wire. Use your own design. Most JA hams use wire construction.
The final configeration in the air should be as close to a vertical rectangle as possible.
This antenna is horizontally polarized.....lay it on it's side for vertical polarity!
Experiment with your favorite support and try to keep conductor size under 1/4 inch.
See the Japan site link below for more info.

6 Meter Version

THE MATH

Calculating the lengths for the Hentenna is simple and straightforward and can be used for HF THRU UHF and possibly beyond.
Just start with 1 meter = 39.36 inches
1 inch = 2.54 cm
1 wavelength = 6 meters = 6 x 39.36 = 236.16 inches

1/2 wavelength per side = 3 x 39.36 =118.08 inches
1/6 wavelength = 6/6 =1 wavelength = 39.36 inches
1/10 wavelength = 236.16 x .1 = 23.61 inches
Some adjustment of lengths may be required for peaking at design frequency. Experiment!

From the above calculations we arrive at the lengths for the Hentenna:

1/2 wavelength sides = 118.08 inches each
1/6 wavelength top, bottom and coax connection element = 39.36 inches
1/10 wavelength matching point = 23.61 inches up from each side of bottom element.

Please note that this is not just a full wave loop. the total length around the outside of it is 314.88 inches which comes out to be 1 1/3 wavelength long. AT first glance the above measurements make this antenna to be HUGE! IT IS NOT. It is only about 9.8 feet tall by 3.28 feet wide! A scaled down Hentenna for 2 meters would be 1/3 it's size or about 39 inches by 13 inches!

EXPERIMENT! EXPERIMENT!

editors comments:

" I personally have not built this antenna YET and cannot warrenty it's performance....make sure you use low power when adjusting the swr and send me any comments you desire about this "strange" Hentenna. IF YOU HAVE EXPERIENCE WITH THIS ANTENNA AND HAVE ANYTHING TO ADD, PLEASE LET ME KNOW."

N4UJW

<<<<<<<<<<<<<<<Make sure to see the WA0IPT 2 METER VERSION

Flash! Just added....6 Meter Hentenna by K5USS HERE!

CLICK HERE TO SEE THE SITE IN JAPAN FROM TAKA JR1LZK AND BE SURE TO NOTE ANY TRANSLATION ERRORS!

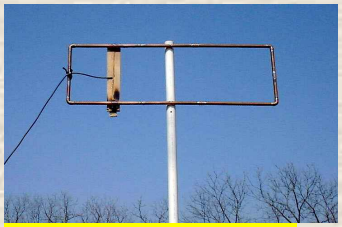
BACK TO ANTENNA LAB

W4RNL L.B.CEBIK ON THE HENTENNA

Why pay full retail for inkjet supplies when you can save between 50%-70%! Click here and start saving!

New 2 Meter Hentenna by Terry, WA0ITP SEE BOTTOM OF PAGE!

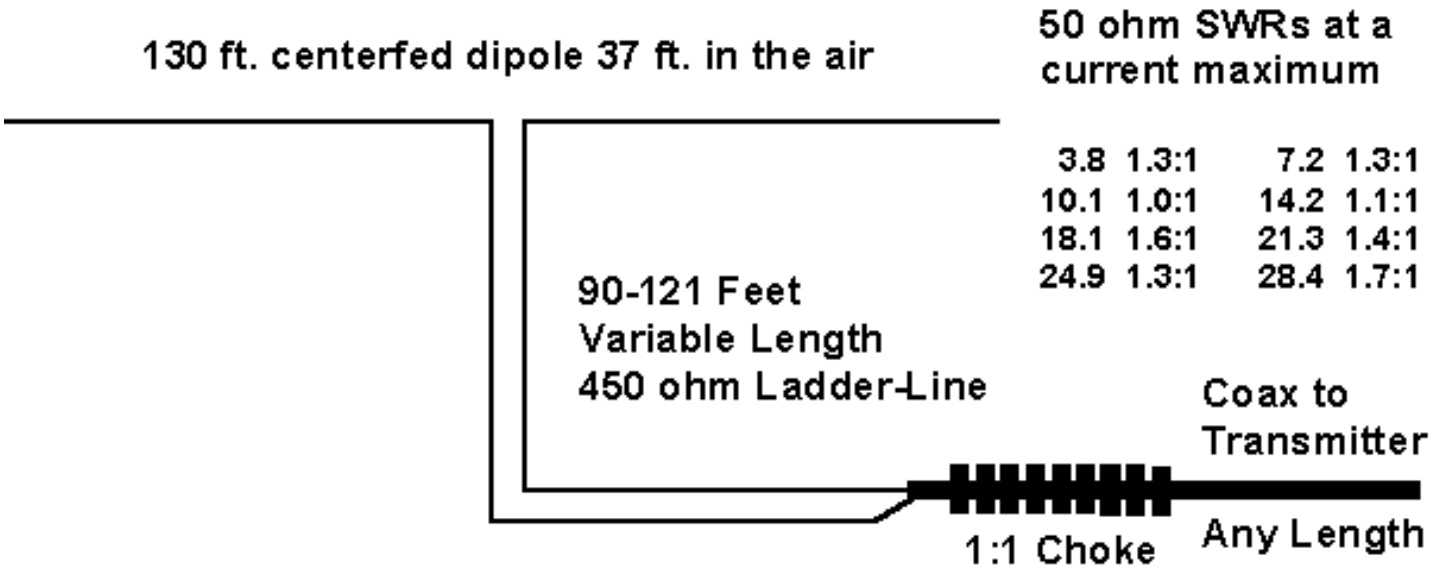
ON the AIR! Click the photo for 2 Meter Hentenna version by Terry, WA0ITP



New
6 Meter Hentenna
by
Charlie
KK5USS
SEE BOTTOM OF PAGE!

W5DXP's No-Tuner, All-HF-Band, Horizontal, Center-Fed Antenna

No-Tuner All-HF-Band Antenna

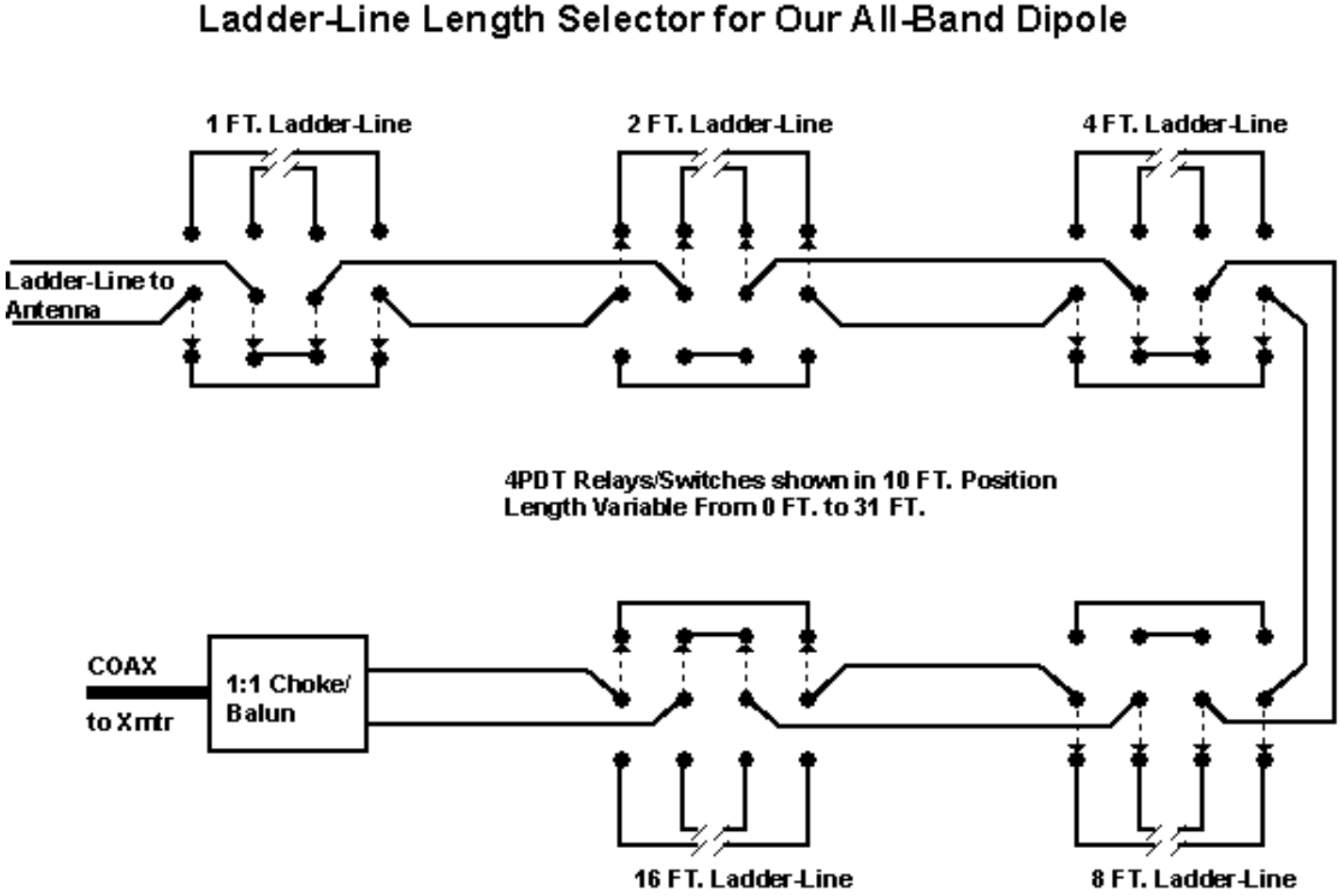


Ferrite beads for the 1:1 choke are available from Amidon Associates
Ten of the FB-77-5621 beads will work for RG-58
Fourteen of the FB-77-1024 beads will work for RG-213
Thirty of the FB-77-6301 beads will work for RG-174 or RG-316 (Teflon)

The No-Tuner, All-HF-Band, Horizontal, Center-Fed Antenna is our old friend, the 80 meter halfwave dipole dressed up a bit. By varying the length of the 450 ohm ladder-line feeding the antenna, we can achieve an SWR of less than 2:1 on all frequencies on all HF bands with the exception of the lowest part of 80m. On 75m, we are feeding the antenna with a half-wavelength of ladder-line. On 40m, we are feeding it with 3/4 wavelength of ladder-line.

No antenna pruning required. My transmission line really does tune my antenna system.

Special thanks to Walt Maxwell, W2DU and Jim Bromley, K7JEB.



The Ladder-Line Length Selector actually does tune the antenna system so no conventional "antenna tuner" is needed - no coils and no capacitors. Switches or relays (remote control) can be used for the switching function and should be sized according to the RF power levels involved. W5DXP presently uses ten DPDT Knife switches attached to a piece of plexiglas mounted in the hamshack window. For portable or backpacking use, the length selector function can be performed simply by 1/2/4/8/16 foot pieces of ladder-line with mating connectors on the end. The proper length of ladder-line is selected to cause resonance in the antenna system.

Here's a table that explains it all. The transmission line always consists of a

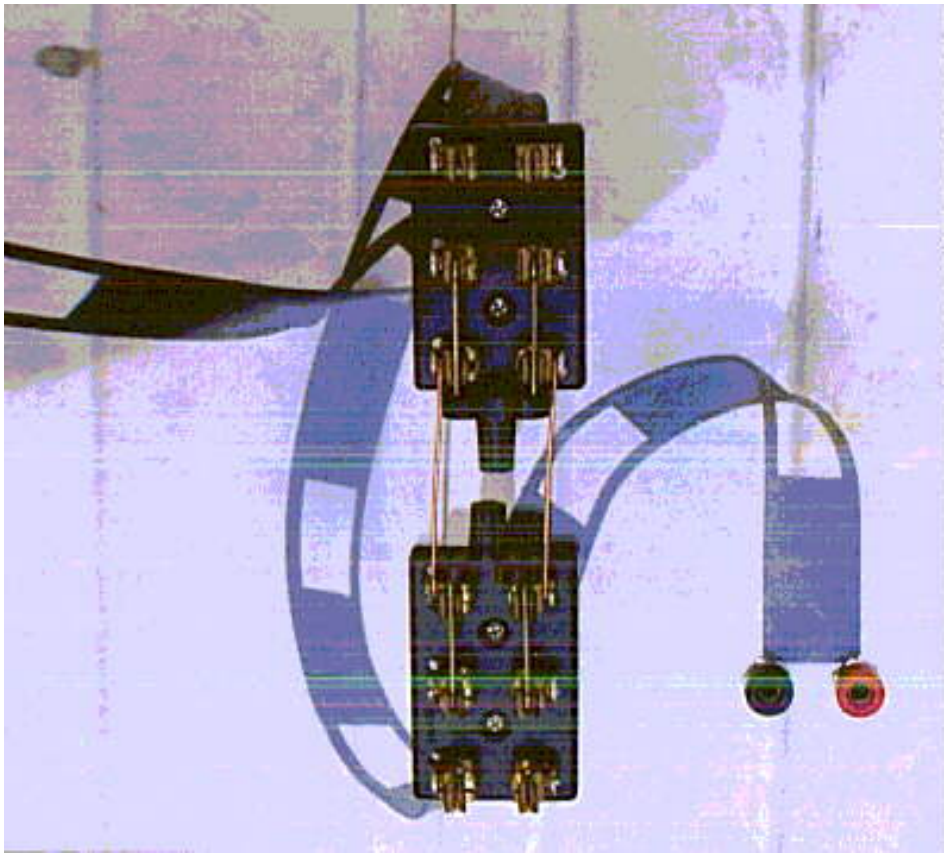
matching section and from zero to six halfwavelengths of ladder-line. The impedance at the antenna is shown along with the 450 ohm SWR and the impedance at the transmitter is shown along with the 50 ohm SWR, i.e. the SWR seen by the transmitter.

..Freq-MHz..	..T-line length = Matching Section + 1/2WL's..	..Impedance at XMTR..	..50 ohm SWR..	..Impedance at Antenna..	..450 ohm SWR..
3.8	109.5' = 109.5' + 0	69 ohms	1.4:1	71+j84	6.6:1
7.2	92.0' = 30.5' + 1x61.5'	40 ohms	1.2:1	4939-j716	11.2:1
10.125	99.4' = 12.0' + 2x43.7'	50 ohms	1.0:1	116-j510	9.1:1
14.2	110.2" = 16.6' + 3x31.2'	53 ohms	1.1:1	2120+j1886	8.5:1
18.14	101.9' = 4.3' + 4x24.4'	81 ohms	1.6:1	111-j267	5.5:1
21.3	94.8' = 11.6' + 4x20.8'	70 ohms	1.4:1	1210+j1378	6.4:1
24.95	94.1' = 5.35' + 5x17.75'	65 ohms	1.3:1	186-j593	6.9:1
28.4	102.8' = 9.2' + 6x15.6'	87 ohms	1.7:1	721+j1009	5.2:1

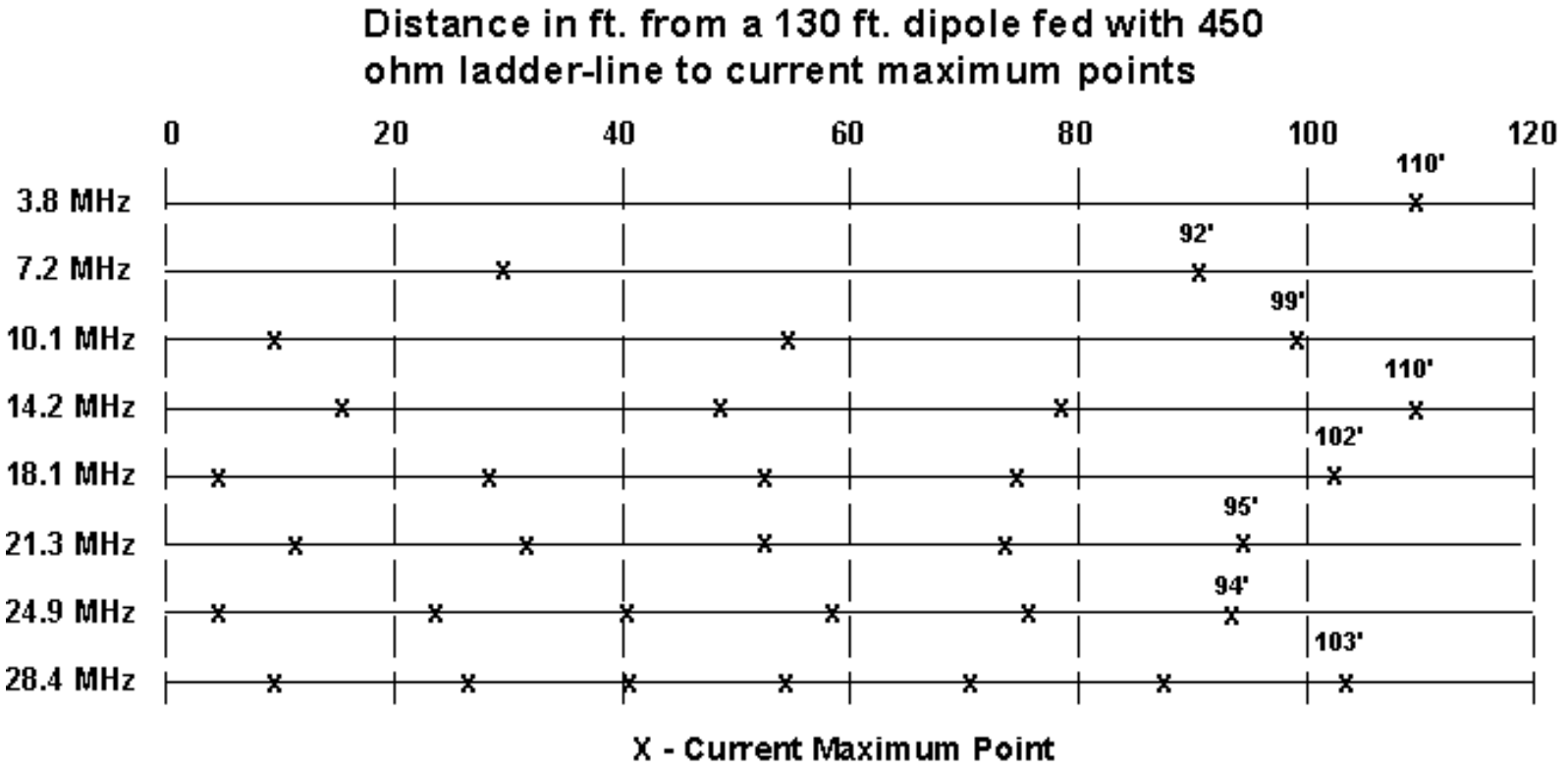
75M Graphs 40M Graphs 30M Graphs 20M Graphs 17M Graphs 15M Graphs 12M Graphs 10M Graphs

---

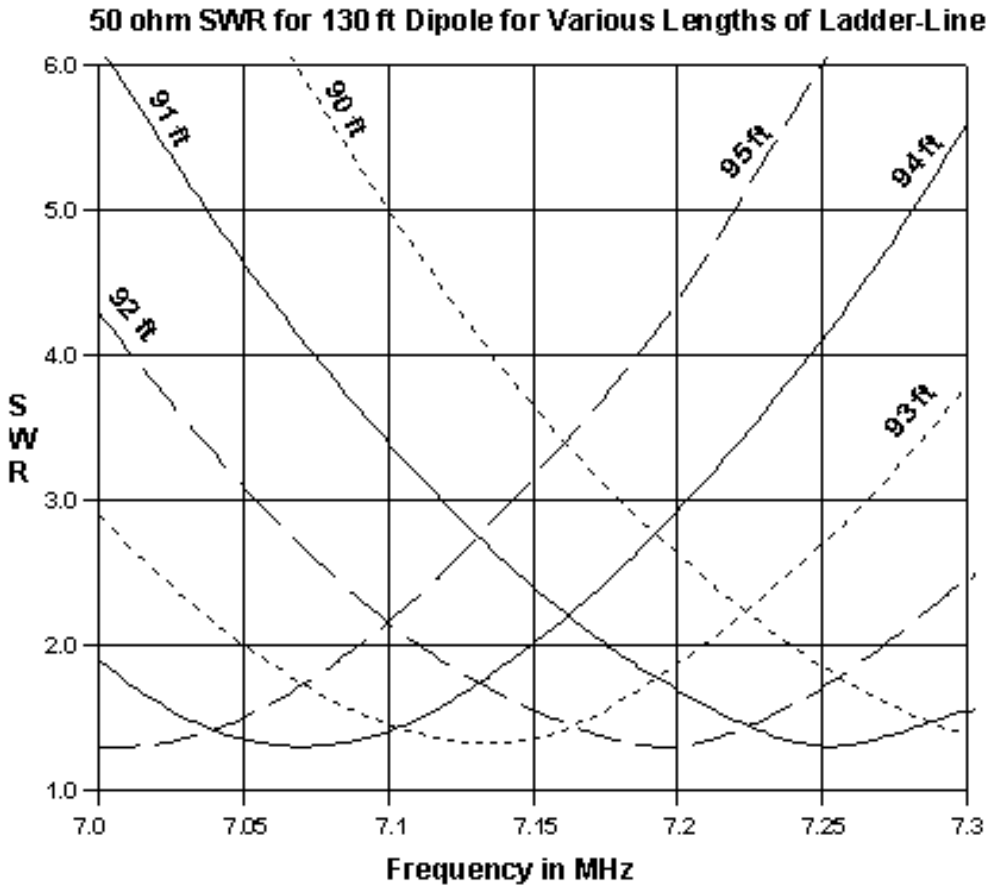
Here are the ten DPDT switches mounted on a piece of plexiglas that mounts in W5DXP's hamshack window. It shows the ten DPDT switches with the one foot, two feet, and four feet loops installed. The eight feet and 16 feet loops are not installed yet in this picture. RF flow is right to left from banana socket set to banana socket set. When installed in the hamshack window, the switches are on the inside and the loops of ladder-line are on the outside.



Here's a close up view of the one foot section. The RF flow is right to left into the banana sockets. The switches are shown in the shorted position, i.e. the one foot loop is floating completely out of the circuit to avoid capacitive effects. The bare copper wires in the center are the short. When the switches are thrown into the other position, the one foot loop is inserted into the circuit and the short is completely out of the circuit. This is the cleanest mechanical configuration W5DXP could think of but there might be a better way.

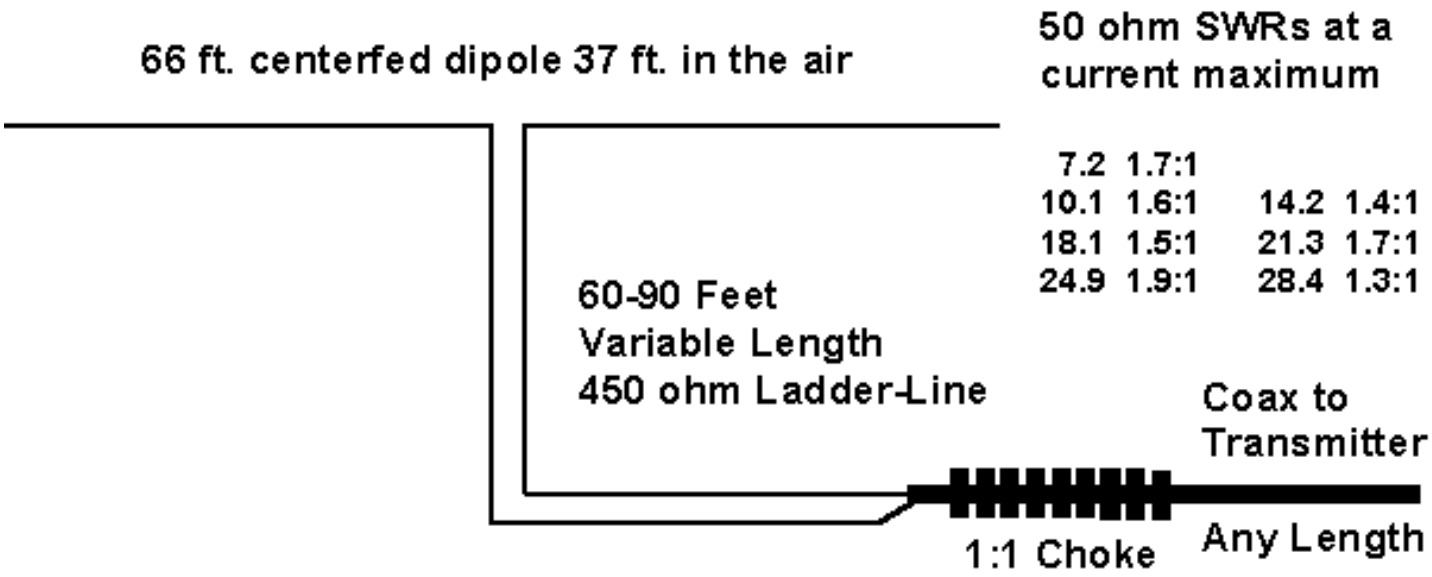


This is a plot of all the current maximum points between the antenna and W5DXP's shack. The transmission line is 90 feet long and the Ladder-Line Length Selector can add in an additional zero to 31 feet for a total of 90 feet to 121 feet. 90 feet matches the antenna on about 7.3 MHz and 121 feet matches the antenna on about 3.6 MHz. The matching points for all the other HF bands lie between these two extremes. Note that if a fixed length of ladder-line needs to be chosen for best results with this antenna, that length should be around 100 ft. which should work with internal autotuners. Caution: Do not expect a similar antenna erected in a different location to exactly match W5DXP's results. The antenna environment has a large effect on the antenna characteristics so W5DXP's results are only approximations when applied to other antenna locations and environments. Mounting this antenna in an inverted-V configuration, for instance, is likely to change the characteristics by an unexpected amount. "450" ohm ladder-line characteristic impedance varies all the way down to 375 ohms for the #14 stranded configuration and velocity factor varies among the different manufacturers and batches of ladder-line.



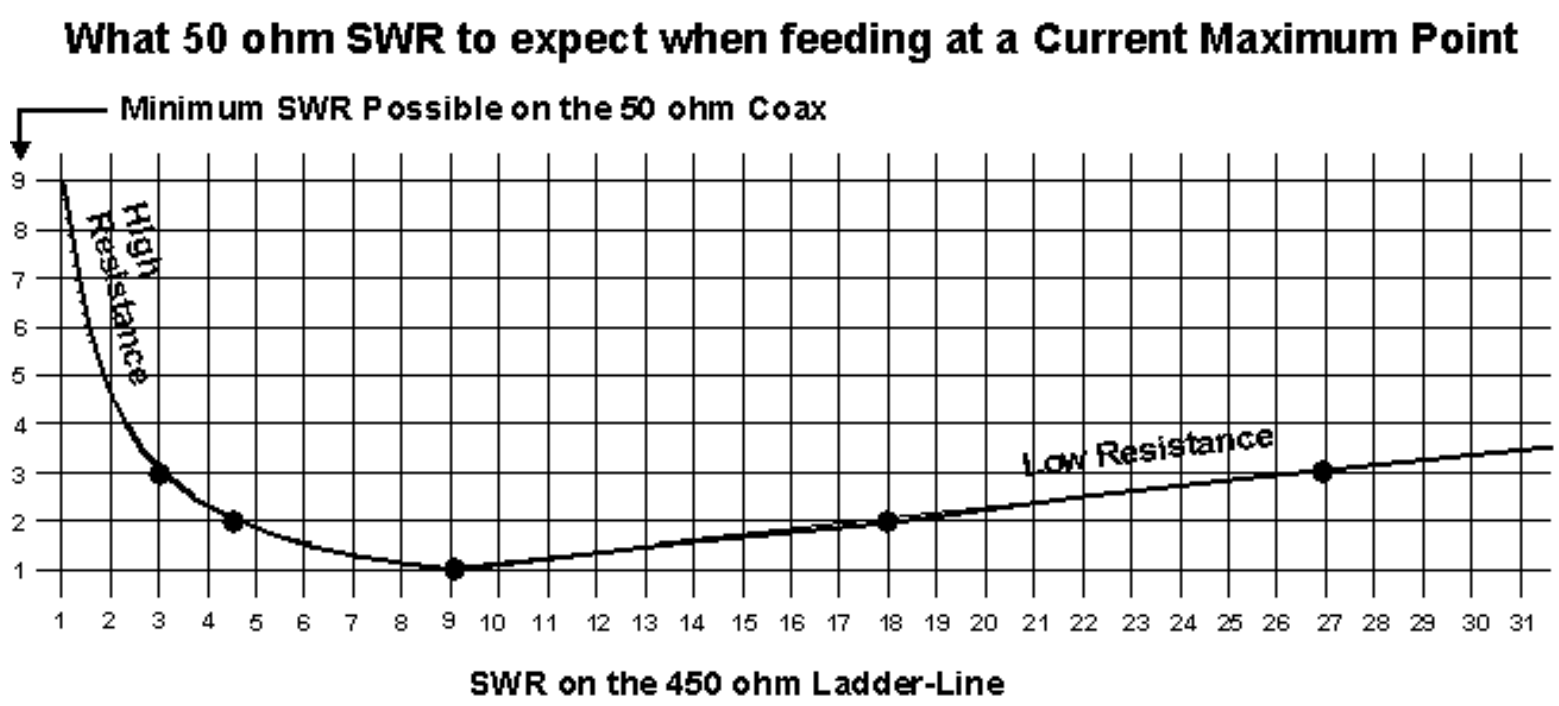
Who says a full-wave dipole is hard to match? **Here's what EZNEC predicts will be the 50 ohm SWR across the 40 meter band for W5DXP's No-Tuner All-HF-Band Antenna given the chart lengths of ladder-line. Similar SWRs occur in similar patterns on the other HF bands.**

No-Tuner "Shorty" HF-Band Antenna



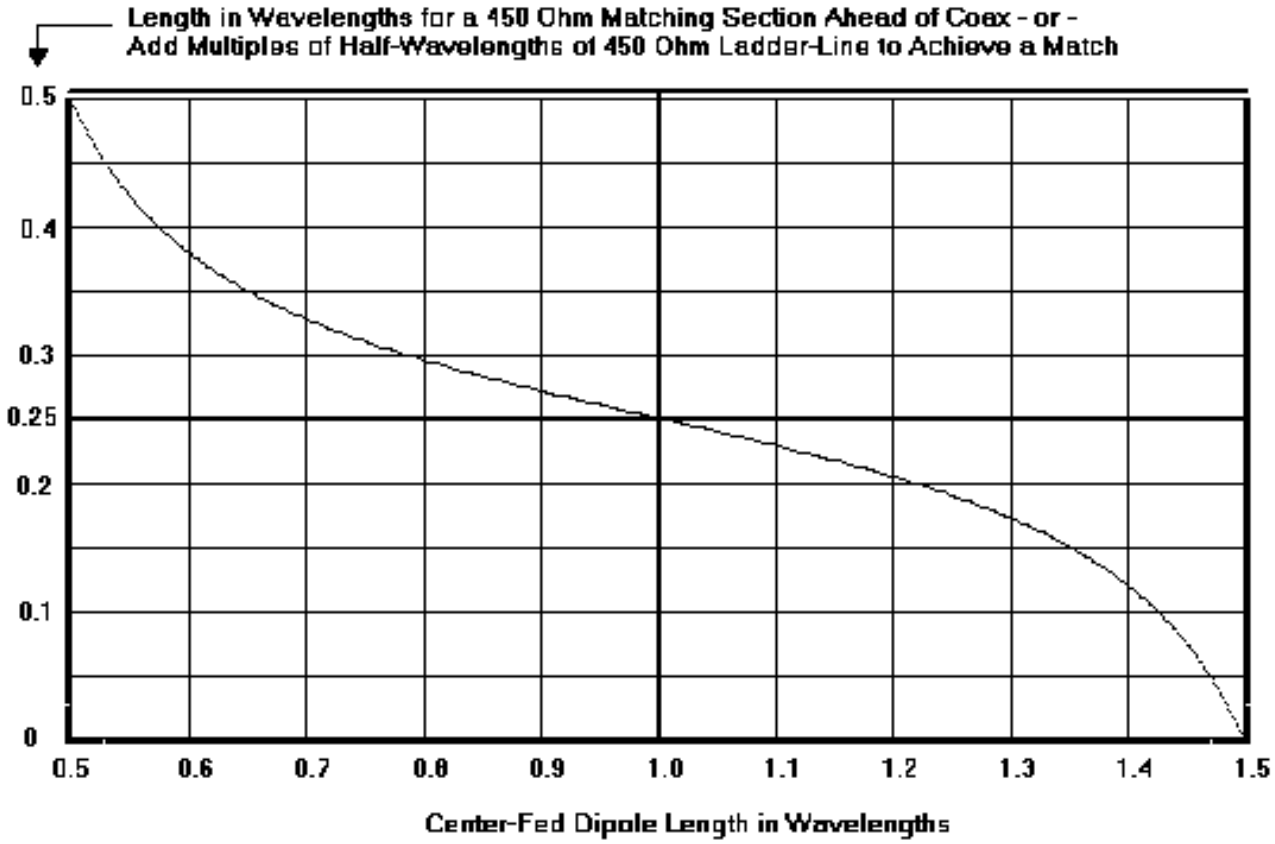
Ferrite beads for the 1:1 choke are available from Amidon Associates
Ten of the FB-77-5621 beads will work for RG-58
Fourteen of the FB-77-1024 beads will work for RG-213
Thirty of the FB-77-6301 beads will work for RG-174 or RG-316 (Teflon)

For those who don't have the space for a 130 foot antenna, here's a "Shorty" version designed to work on all HF ham frequencies above 7 MHz. Like the bigger version, the 50 ohm SWRs predicted by EZNEC are below 2:1 for the bands of interest. This antenna will work on 75 meters at reduced efficiency with a matching network or tuner.



Here is the physics that makes it all possible. Any 450 ohm SWR between 4.5:1 and 18:1 will result in a 50 ohm SWR of less than 2:1 IF the antenna system is fed at a current maximum point. ***Moral: Make your center-fed HF antenna system at least a half- wavelength long at your lowest operating frequency and feed it at a current maximum point on the ladder-line.***

Optimum Length For A Matching Section



This graph shows the optimum length for a matching section when feeding a center-fed horizontal dipole. The bottom of the chart is normalized to wavelengths so it works for most HF frequencies and most popular lengths of center-fed wire dipoles. The left side of the chart indicates the optimum wavelength for a 450 ohm ladder-line matching section for connection to coax or connection to a multiple of half-wavelengths of 450 ohm ladder-line.

Example: Assume a 102 ft dipole on 7.2 MHz. 102/(936/7.2) equals 0.785 wavelengths on 7.2 MHz. Reading the matching section length from the graph yields 0.3 wavelength. A wavelength of 450 ohm ladder-line on 7.2 MHz is 886/7.2= 123 ft. 0.3 times 123 equals 36.9 ft for the 7.2 MHz matching section. Add 123/2 = 61.5 ft if 36.9 ft is too short for a total of 98.4 ft.

The following BASIC program approximates the optimum feedline lengths given the length of a horizontal dipole and the frequency. It works for both 300 ohm and 450 ohm ladder-line by assuming a velocity factor of 0.8 for the 300 ohm and 0.9 for the 450 ohm. The results are only approximations based on EZNEC and must be fine-tuned to perfection in reality. Cut and paste this program to Notepad and store it in the BASIC directory as Imax.bas

Note: This BASIC program only works for *horizontal* dipoles, not for inverted-V's or any other folded antenna.

```
10 REM This program calculates optimum ladder-line
20 REM lengths given dipole length and frequency
30 CLS
40 INPUT "Enter Frequency in MHz ", freq
50 INPUT "Enter Dipole Length in Feet ", diplenft
60 INPUT "Enter either 450 or 300 for Z0 ", Z0
70 IF Z0 = 450 THEN LLWL = 886
80 IF Z0 = 300 THEN LLWL = 787
90 dipwl = diplenft / (936 / freq)
100 IF dipwl < .5 THEN dipwl = dipwl + 1
110 IF dipwl < 1.5 THEN GOTO 140
120 IF dipwl > 1.5 THEN dipwl = dipwl - 1
130 GOTO 110
140 fedlinwl = .25 - (TAN(2.5 * (dipwl - 1))) / 12.02
150 fedlinft(0) = (LLWL / freq) * fedlinwl
160 FOR i = 1 TO 7
170 fedlinft(i) = fedlinft(0) + i * ((LLWL / 2) / freq)
180 NEXT i
190 PRINT "Imax points (Current Loops) at"
200 FOR i = 0 TO 7: PRINT fedlinft(i), : NEXT i
210 PRINT : PRINT : GOTO 40
220 END
```

Homepage DK7ZB

DK7ZB

Martin Steyer, Die Aue 2, D-37269
Eschwege-Germany

new e-mail: dk7zb@fox28.de

Very important notice: e-mail adresses like CALL@DARC.DE are generated very easily for automatically working spammers. I get about 15-25 spam-mails per day. If I cannot read out my box for some days (QRL, holiday, e.g.) it is possible, that mails will be erased because the box is full or they will be found between the spam-mails, which will be erased too after some days.

If you do not get an answer within some days, please send another e-mail. Please understand my problems and use only my new adress.

This homepage is dedicated to the homebrewing of equipment, especially for building HF- and VHF-Yagi-antennas.

Sorry, a lot of pages in German language, because DK7ZB is writer for some ham-magazines in Germany.

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Last Update: 07-July-2004

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50MHz UKSMG

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New Design and Matching of Yagi-Antennas for HF and VHF/UHF

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The Gain of the Axial-Mode Helix Antenna

by Dr. D.T. Emerson, National Radio Astronomy Observatory (NRAO++).
Antenna Compendium Volume 4, pp 64-68, 1995, published by the ARRL.

(Copyright on original article is held by the ARRL)

Summary

The axial-mode helix antenna, first described by Kraus in 1947, is probably the most widely used circularly-polarized antenna, either in space or on the ground. There are conflicting claims for the gain of the antenna; most amateur literature, and even many standard textbooks, quote gains which are far too optimistic. More realistic gain relationships are available now in the professional antenna journals, but are not well known in amateur circles.

This article summarizes probably the most extensive numerical modelling calculations on the helical antenna ever performed. NEC-2 was used to model some 10,000 different helical antennas, systematically changing the physical parameters of the antenna and investigating the effect on gain and feed match. For each calculation, between 600 and 1400 segments were used to model an antenna; with frequent checks on the validity of the calculations. Some 3000 hours of networked Sparc workstation compute cycles were used for the study.

The modelling data were compared with results from the professional antenna literature. Reassuringly, the modelling gives results which are intermediate between published experimental and theoretical work. The maximum possible gains are up to 4 or 5 dB lower than those derived from the original Kraus formula for gain. The maximum gain increases much more slowly with increasing antenna length than the simple Kraus formula would predict.

An empirical expression for the maximum possible gain Gmax of the helical antenna as a function of its length L in wavelengths is:

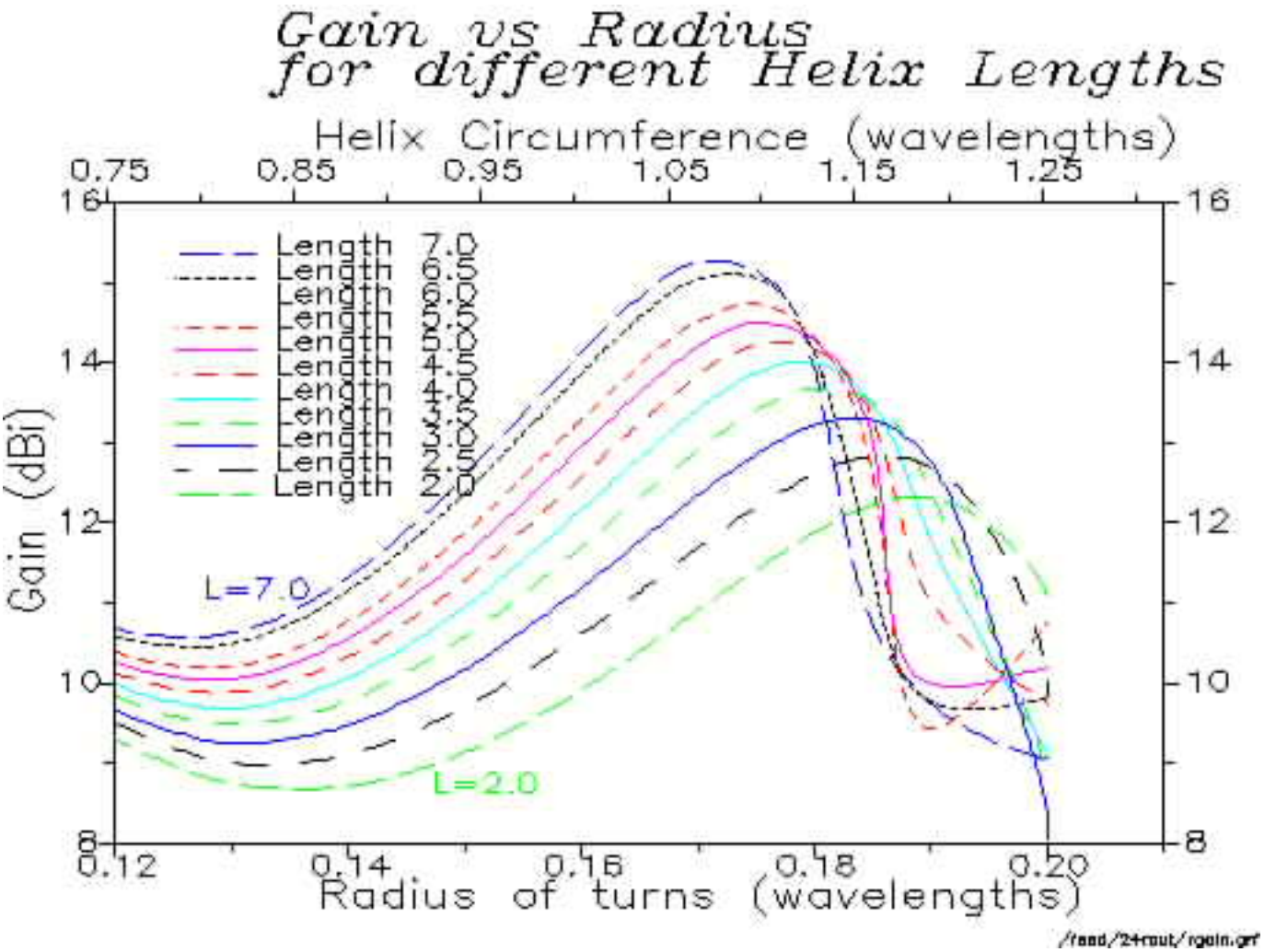
```
Gmax(dB) = 10.25 + 1.22 L - 0.0726 L^2
```

This expression is only valid for lengths L between 2 and 7 wavelengths.

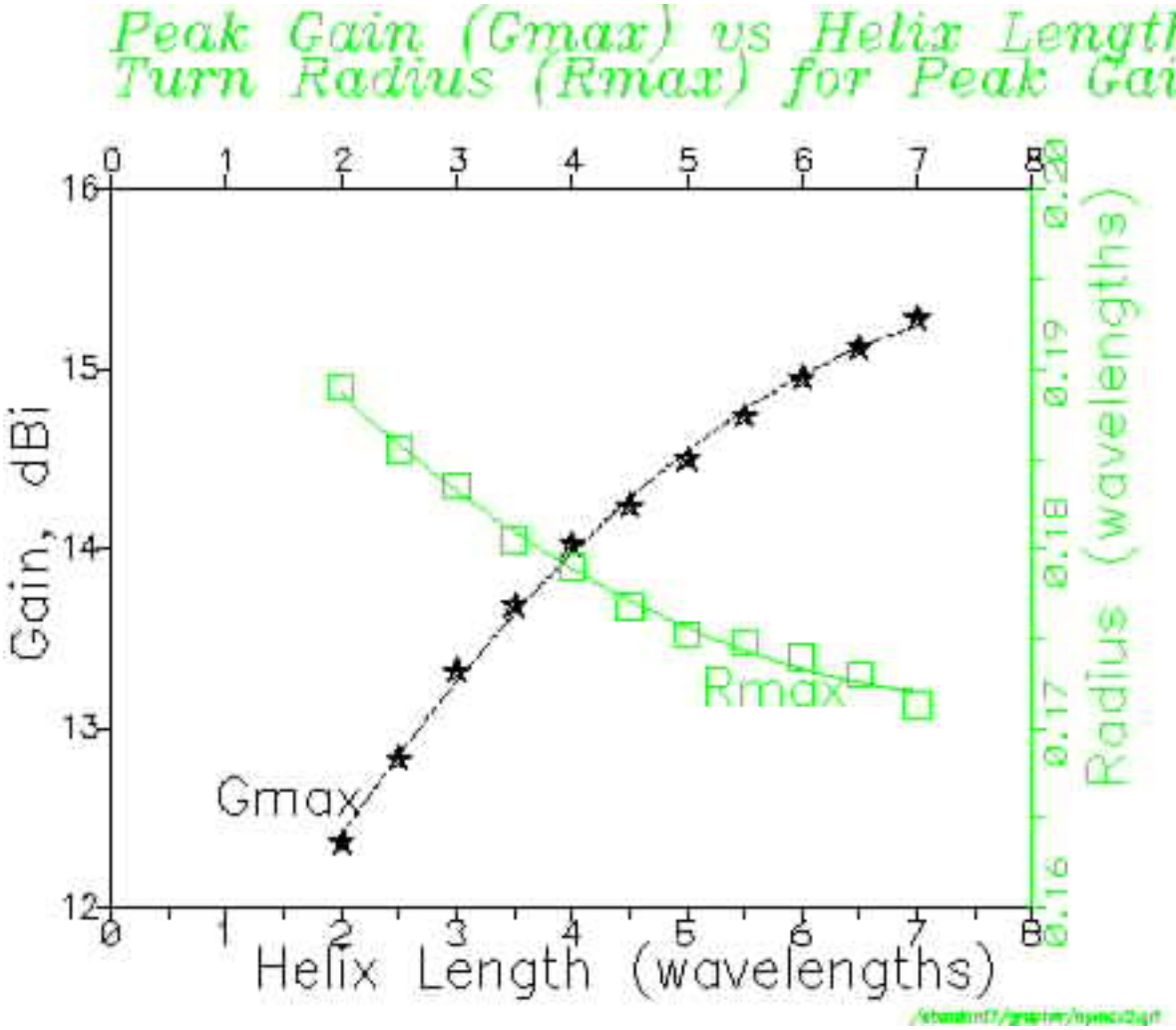
An empirical expression for the turn radius Rmax at which peak gain occurs as a function of length L in wavelengths is:

```
Rmax = 0.2025 - 0.0079 L + 0.000515 L^2
```

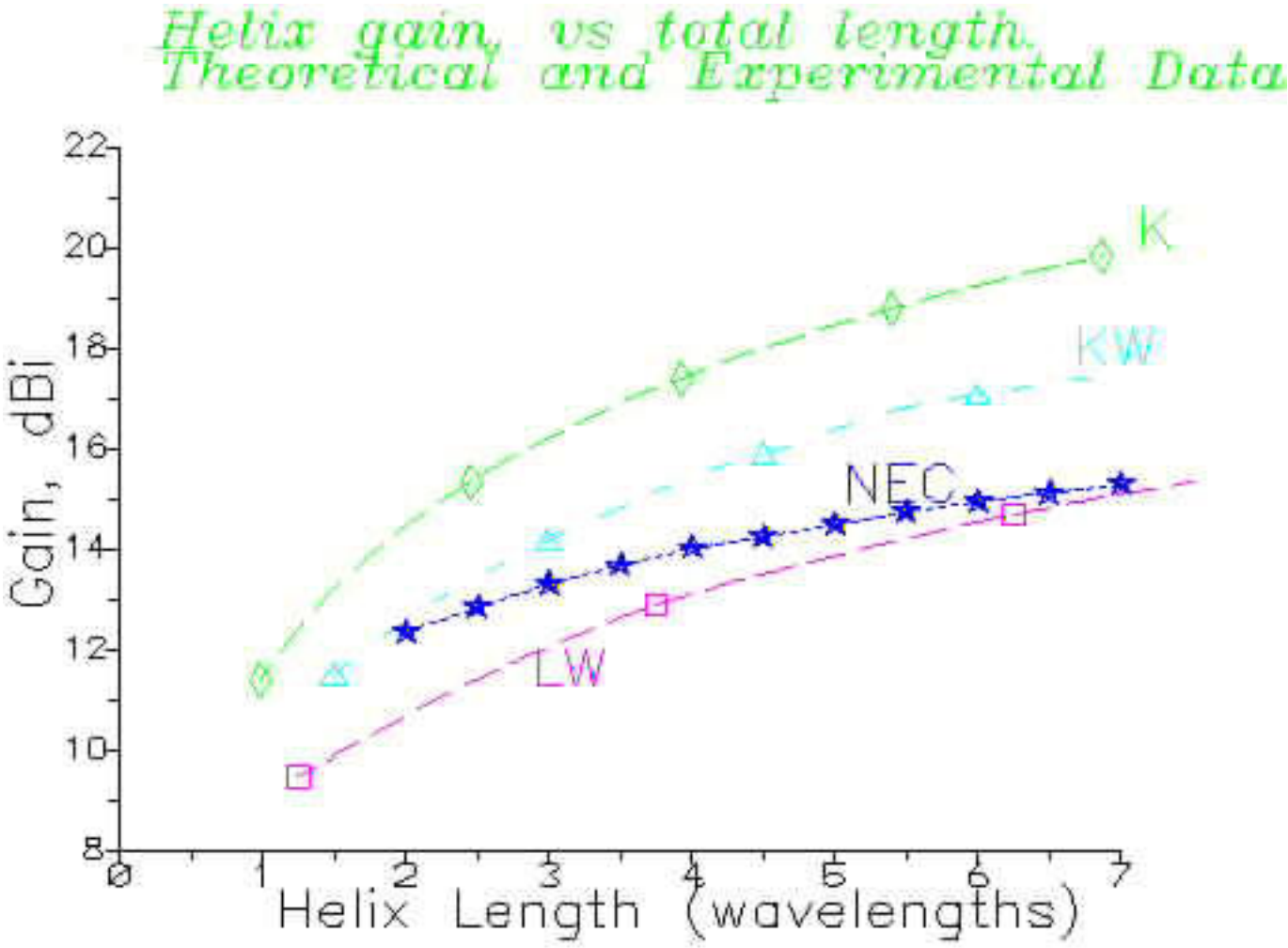
Again, this is only valid for lengths between 2 and 7 wavelengths.



Gain vs length and turn radius, with turn spacing 0.24 lambda. Note that the radius at which peak gain occurs decreases with antenna length.



The maximum realizable gain (stars, left axis) as a function of helix length, and the turn radius (squares, right axis) at which peak gain occurs.



The peak gain of a helix as a function of length, comparing NEC modelling, experiment and theory. K is the gain from the Kraus formula. KW (long dashes) are the measurements of King and Wong. NEC (short dashes) is from the numerical modelling in this article, and LW is from the theory of Lee and Wong. The modelling results are intermediate between theory and measurement.

In summary, this extensive modelling study is in good agreement with theory and measurement from the professional literature, all of which show that the simple Kraus formula for gain of a helix is far too optimistic - by up to 4 or 5 dB.

The modelling shows that, at a given value of turn spacing, the optimum turn radius for peak gain decreases slightly as the helix is made longer. The gain is almost independent of wire diameter, or of the presence of a short feed stub between the ground plane and the start of the helix. The resistance of wire used to construct the helix, even if several times worse than aluminium, has little effect on efficiency. A half-lambda square groundplane is nearly as good as an infinite groundplane. The use of radials, rather than a continuous groundplane, gives a gain penalty of some 3.5 dB.

Please consult the original article for more results, and for literature references. A preprint of a more

detailed article on the gain of the helix antenna, prepared for the professional literature, is available from the author. (Please send e-mail to demerson@nrao.edu.)

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E-mail (demerson@nrao.edu)

Horizontal Half-wave Dipole above a Counterpoise

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Introduction

All antennas that are situated close to the ground are affected by that ground to some extent.

The most obvious effect is that the ground forces the antenna's radiation pattern to appear in the half-space above the ground. This is illustrated by comparing the radiation around a monopole fed against ground to that of a dipole in free-space. The monopole has twice the power in the hemisphere above the ground compared with the power in either hemisphere of symmetry for the dipole in free-space. The nature of this reflection of energy above the ground is governed by the polarisation of the antenna and the ground effectiveness. The dielectric constant and conductivity of the ground determines how well the ground acts as a conductor and hence a reflector.

A less obvious effect is that the ground absorbs energy from the antenna; this energy is wasted in the ground's intrinsic resistance (Ref 3).

The placement of an artificial ground, or counterpoise, can decrease the ground losses and enhance the performance of a horizontal antenna. The enhancement can be investigated by performing comparative measurements of the feedpoint resistance or by performing computer simulations.

This article is based on computer simulations using NEC-2 (Ref 1) for antenna systems at 1.825 MHz over average ground with a relative dielectric constant of 13 and conductivity of 5 milli-Seimens per metre [13,5]. The simulations were performed for ideal (lossless) antenna elements with a diameter of 1.22 milli-metres.

Terminology

Displacement is the term used for the distance between the counterpoise (which is the antenna system's lowest element) and the ground. Separation is the term used for the distance between the driven-element and the counterpoise. Displacement, in the regular dipole case, is the distance between the dipole and the ground.

Results

Figure 1 illustrates the efficiency of the horizontal half-wave dipole at various displacements. In general, the efficiency improves as the dipole is raised higher. Note that 100 percent efficiency is not achieved at a quarter wavelength displacement because, even at this height, the ground has introduced some losses into the antenna system.

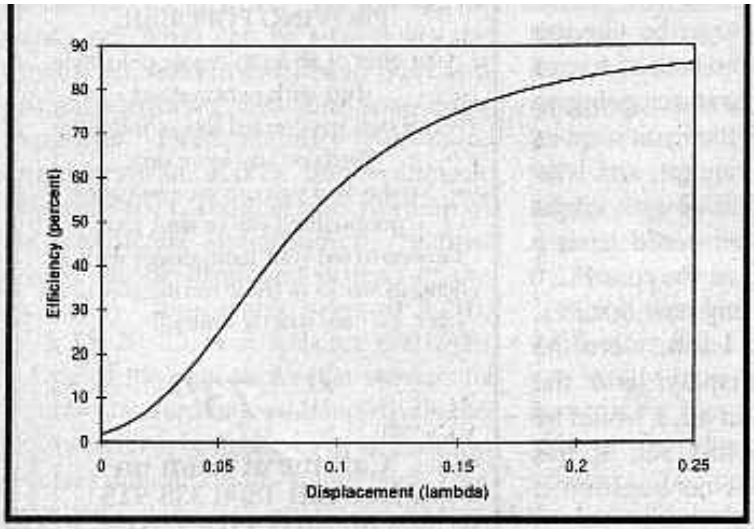

Figure 1 – Efficiency of a half-wave dipole above ground [13,5].

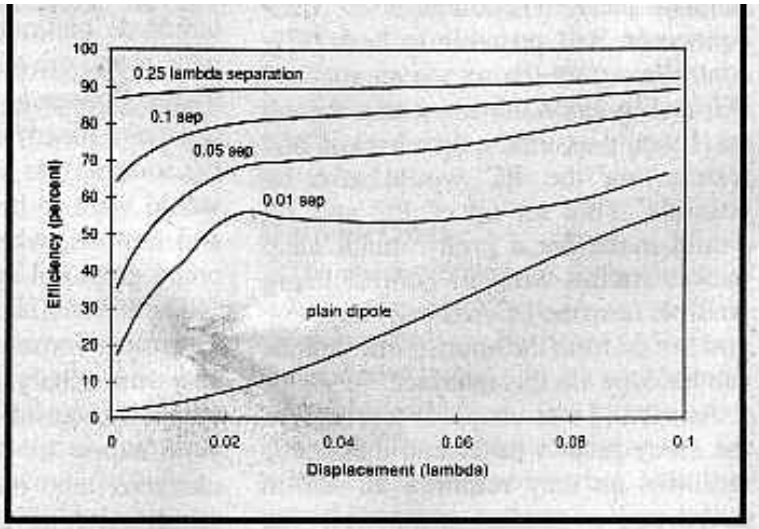

Figure 2 – Efficiency of a half-wave dipole separated from a counterpoise with various displacements above real ground.

Figure 2.0 illustrates the efficiency for various counterpoise systems and a dipole above ground. Notice that there is a knee in the counterpoise curves, which indicates that there is an optimum displacement. Be careful when interpreting the graphs, increasing the separation suffers from the law of diminishing returns; the efficiency of a counterpoise system must be compared with that of a dipole at the same equivalent height, ie. the counterpoise displacement plus the driven element separation. Observe how a counterpoise improves the efficiency of a horizontal antenna system.

Figure 3 shows the radiation pattern of a horizontal dipole displaced 0.07 wavelengths above ground.

Figure 4 shows the radiation pattern of a dipole offset by 0.05 wavelengths above a counterpoise that is displaced 0.02 wavelengths above the ground (same equivalent height as the dipole only system).

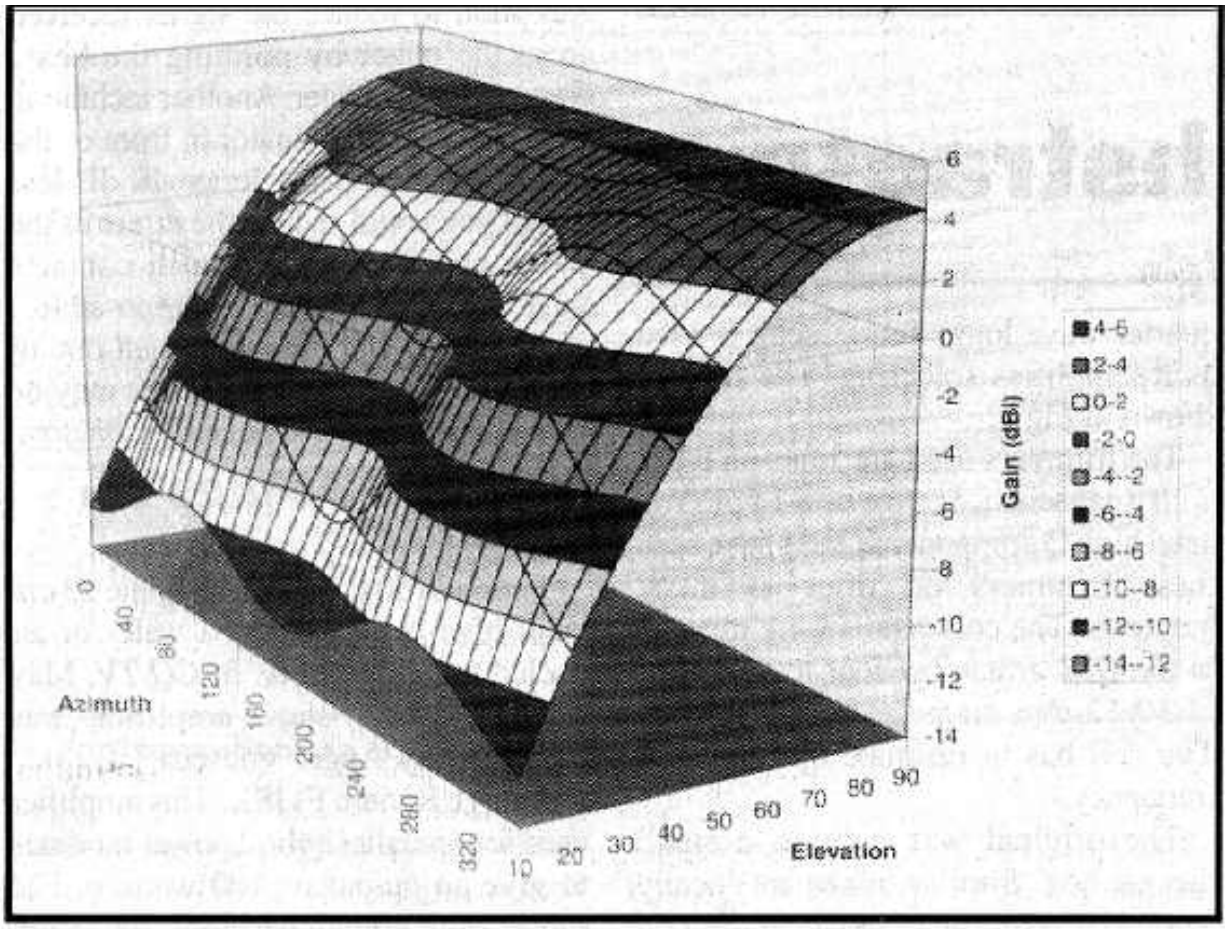

Figure 3 – Radiation pattern from a half-wave horizontal dipole 0.07 wavelengths above ground [13,5].

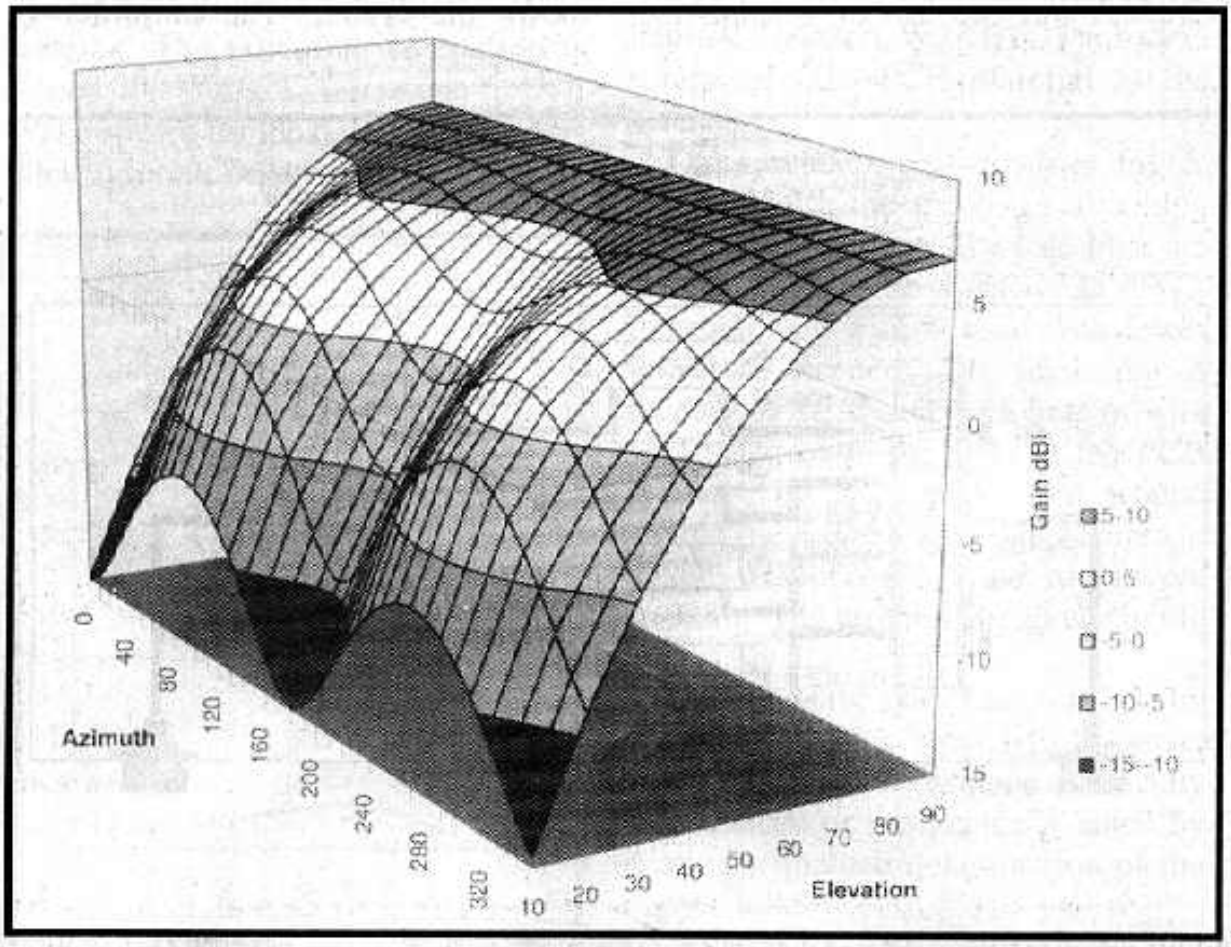

Figure 4 – Radiation pattern from horizontal half-wave dipole above a half-wave counterpoise 0.02 wavelengths above ground [13,5].

The horizontal dipole system has a minimum gain of 0 dBi at 51 degrees elevation, while the counterpoise system has a minimum gain of 0 dBi at about 42 degrees of elevation. Alternatively, antenna gain comparisons at all elevation angles show that the counterpoise gain is greater than the dipole system by 2.53 (0.05 dB. This demonstrates that the counterpoise is a more effective ground system; the counterpoise reflects more energy into the half-space so less is wasted in the intrinsic ground resistance and more is radiated.

Conclusions

The placement of a counterpoise below a horizontal antenna can improve the antenna efficiency by a reasonable amount. The effect under a horizontal antenna is similar to the effect of elevated ground-planes for vertical antennas (Ref 2). You can measure this effect by observing an increase in field strength or by observing the lowering of your feedpoint resistance when adding the counterpoise (Ref 3).

At 160m, with ground parameters [13,5], the effect peaks with counterpoise displacements of about 0.02 wavelengths above the ground.

Increasing the antenna system efficiency with a counterpoise increases the gain at all angles, effectively lowering the effective radiation angles.

References 1.

1. **Computer program NEC-2**, G. J. Bourke, Lawrence Livermore National Laboratory, 1984.

2. **Short Vertical Antennas and Ground Systems**, Ralph Holland, *Amateur Radio*, October 95.

3. **Horizontal Antennas above Real Ground**, Ralph Holland, *Amateur Radio*, October 1996 .

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Horizontal Antennas above Real Ground

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Introduction

Antennas are influenced by the effect of the ground and by the type of conductors from which they are constructed. The feedpoint impedance is the summation of the radiator's self impedance, the mutual impedance of its image in the ground, and the loss resistance.

The loss resistance is the summation of the R.F. resistance in the conductor, and the resistance introduced by consumption of power in ground losses and other media close to the antenna. The conductor's resistance is modified by the skin-effect which causes the current to only flow in the outer parts, or skin, of the conductor. The effect causes the resulting resistance to increases in proportion to the square root of the frequency (see Table 1 and Table 2).

Horizontal antennas are subjected to the influence of a broadsize image in the ground. The antenna and its image are in anti-phase, so radiation tends to be cancelled at low angles and the radiation resistance is lowered because the mutual impedance of the image is subtracted from the self-impedance of the driven element.

Modelling

To quantify the effects of locating antennas above real ground, I have once again resorted to computer modelling using NEC-2 (Ref 1). All simulation results have been performed with 1.22mm diameter wire (SWG #18) and assume lossless conductors.

The simulation results are displayed graphically so you can determine the trends and evaluate your own antennas. The soil parameters for each simulation are enclosed in square brackets. For example, [13,5] represents ground with a relative dielectric constant of 13 and conductivity of 5 milli-Siemens / m (2S = 1/2ohm, while 4S = 1/4 ohm). The selected values are: [5,1] for poor soil, [13,5] for average clay soil, [20,30.3] for good soil and [80,5000] for sea water, which is very close to perfect.

Table 1 and Table 2 are included so you can evaluate conductor resistance losses due to the skin effect (Ref 2).

Results

Figure 1 shows the effects of various types of grounds on a 1.825 MHz horizontal 0.5 wave dipole between 0.01 and 0.25 wavelengths above the ground. Note how over poorer soils the feedpoint impedance is dramatically higher than the resistance for perfect ground. Notice how the feedpoint resistance for a horizontal antenna becomes very low as the antenna approaches a perfect earth. (The feedpoint resistance of a perfect conductor over a perfectly conducting ground is the radiation resistance of the antenna.)

Figure 2 illustrates the overall antenna efficiency; a measure of how much power is radiated over the hemisphere, compared to power fed into the antenna (the missing power is absorbed by the ground).

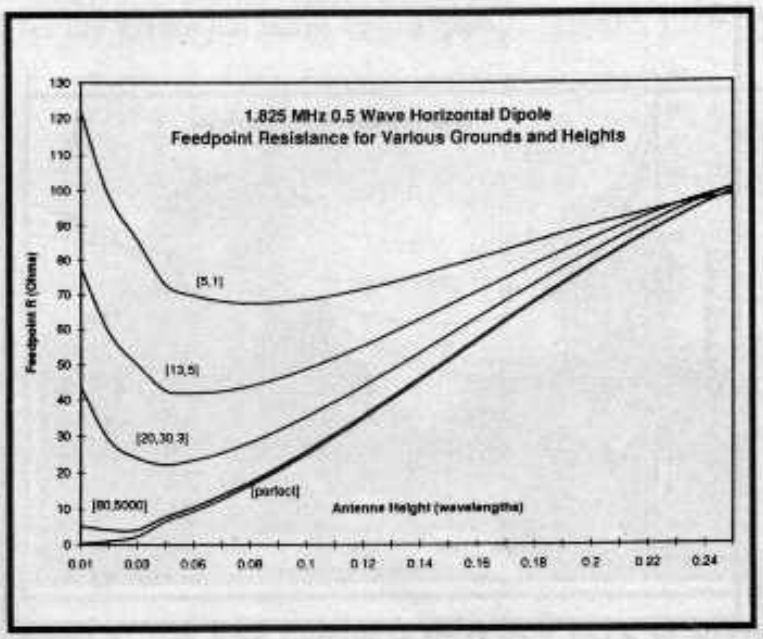

Figure 1

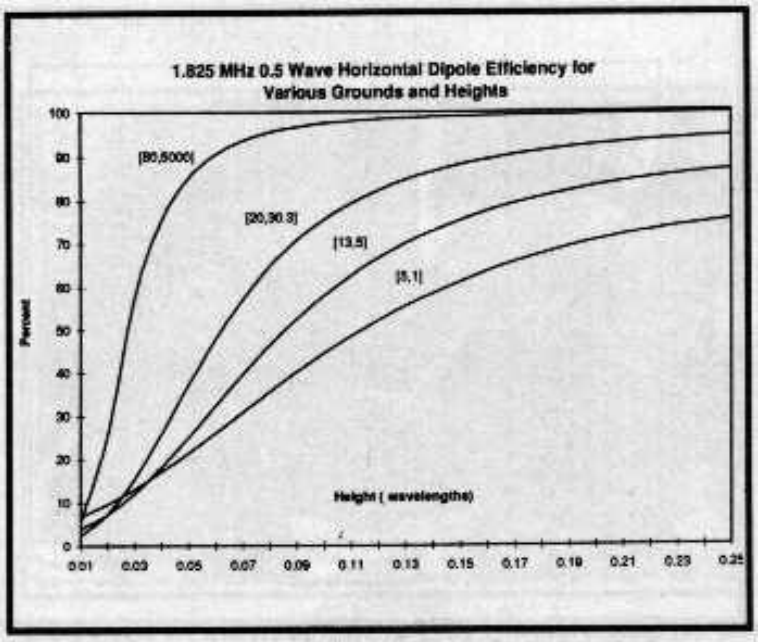

Figure 2

Figure 3 illustrates the effect upon the maximum gain. (However, at 160m, poor ground means the maximum gain is at an elevation of 90 degrees - ie straight up!)

Figures 4, 5, and 6 show the effect upon feedpoint resistance at 3.5 MHz, 7.0 MHz and 14.0 MHz respectively.

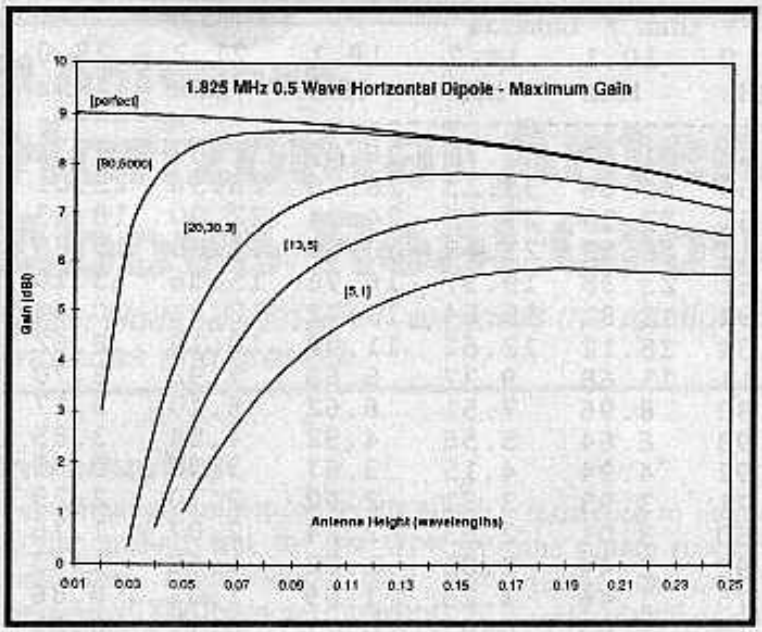

Figure 3

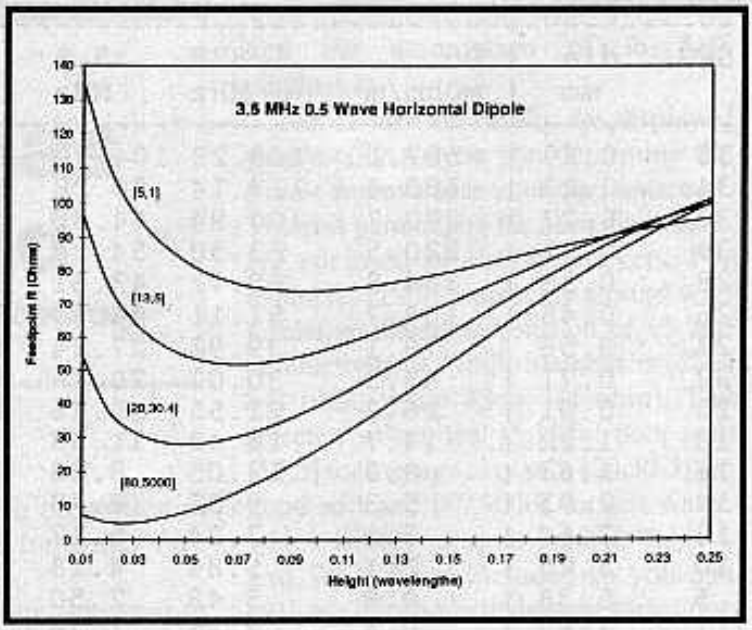

Figure 4

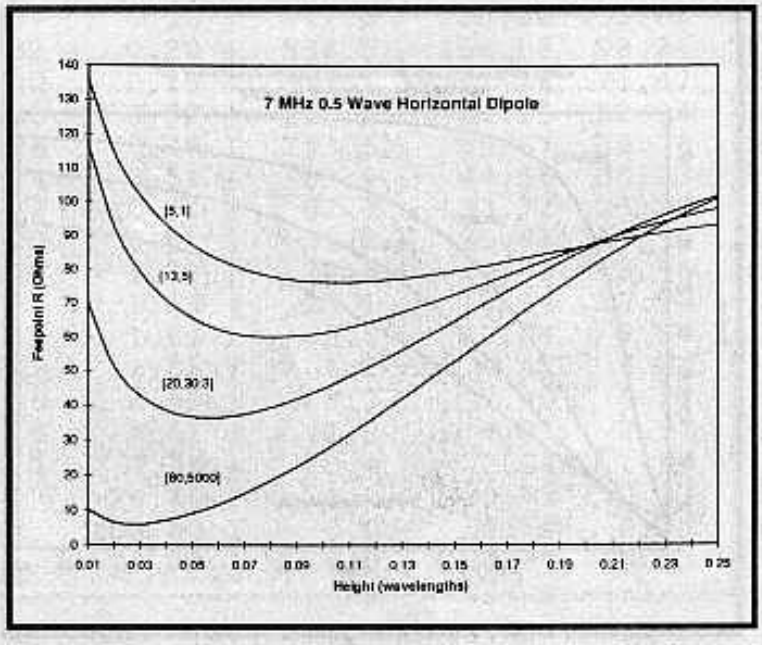

Figure 5

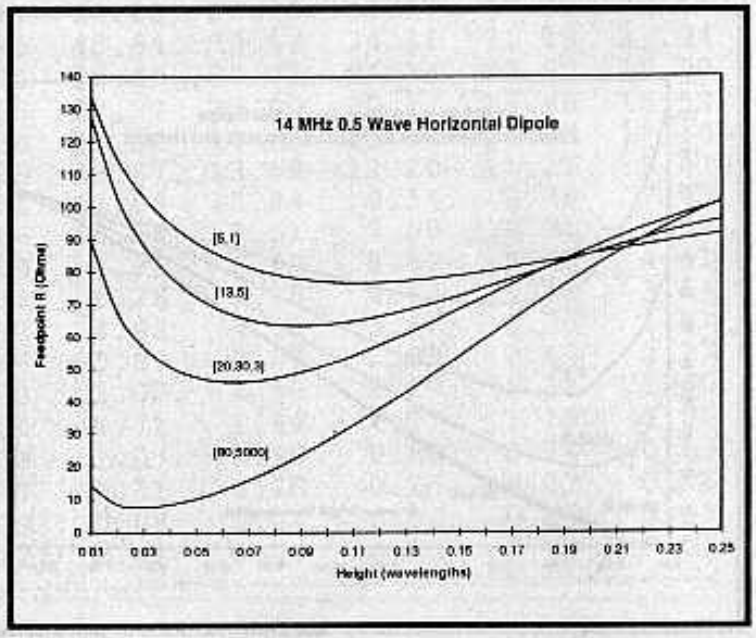

Figure 6

Table 1 and Table 2 give the R.F. resistance of round copper wire at various frequencies. The values are listed in ohms per wavelength. You must halve these values for wires carrying cosinusoidal currents. The resulting value when added to the graphical results accounts for losses in a non-ideal conductor.

Conclusions

The radiation resistance of a horizontal antenna is lowered as the antenna is brought closer to the ground because self and mutual impedances subtract.

As a horizontal antenna approaches lossy media the feedpoint resistance rises due to increasing power losses in the media. You must be wary of this tendency for low antennas to apparently present a good feedpoint resistance.

Poor conductors, or inappropriate conductor sizes, will also introduce loss resistance. This effect is particularly noticeable in cases such as loading coils.

By comparison though, a horizontal radiator has less ground loss than a vertical antenna mounted at the same average height (compare the graphs for horizontals against those for verticals from Ref 3.0).

References

1. Computer program NEC-2, G.J.Bourke, Lawrence Livermore National Laboratory,1984.
2. Fields and Waves in Communication Electronics, Simon Ramo, John R Whinnery and Theodore Van Duzer. Publisher John Wiley and Sons.
3. Short Vertical Antennas and Ground Systems, Ralph Holland, *Amateur Radio*, October 95.

		Ohm / Lambda							
B&S dia	Dc.	1.825	3.5	7.0	10.1	14.2	18.1	21.2	29.0
AWG mm	mOhm/m	MHz	MHz	MHz	MHz	MHz	MHz	MHz	MHz
36 0.13	1356.4	329.59	195.37	114.92	89.43	71.08	60.62	54.77	44.84
34 0.16	841.0	223.82	135.88	82.17	64.82	52.16	44.84	40.72	33.64
32 0.20	536.5	158.03	98.24	60.95	48.64	39.54	34.21	31.19	25.94
30 0.25	337.1	112.16	71.47	45.44	36.65	30.05	26.14	23.90	19.99

28 0.32	212.2	81.15	52.94	34.38	27.97	23.09	20.17	18.48	15.52
26 0.40	133.5	59.67	39.76	26.28	21.52	17.85	15.64	14.36	12.10
24 0.51	83.9	44.56	30.23	20.26	16.67	13.89	12.20	11.21	9.47
22 0.64	52.8	33.72	23.20	15.71	12.98	10.84	9.54	8.78	7.43
20 0.81	33.2	25.82	17.96	12.26	10.16	8.51	7.50	6.91	5.85
18 1.02	20.9	19.89	13.96	9.59	7.96	6.68	5.89	5.43	4.61
16 1.29	13.1	15.44	10.90	7.52	6.26	5.26	4.64	4.28	3.64
14 1.63	8.3	12.04	8.54	5.92	4.93	4.15	3.66	3.38	2.87
12 2.05	5.2	9.42	6.71	4.66	3.89	3.27	2.89	2.67	2.27
10 2.59	3.3	7.40	5.28	3.68	3.07	2.59	2.29	2.11	1.80
5 4.62	1.0	4.07	2.92	2.04	1.71	1.44	1.27	1.18	1.00
1 7.35	0.4	2.54	1.83	1.28	1.07	0.90	0.80	0.74	0.63
2 5.40	0.0	0.73	0.53	0.37	0.31	0.26	0.23	0.21	0.18
100.00	0.0	0.18	0.13	0.09	0.08	0.07	0.06	0.05	0.05
500.00	0.0	0.04	0.03	0.02	0.02	0.01	0.01	0.01	0.01

Table 1 AWG round copper wire resistance.

	Ohm / Lambda							
SWG dia mm	Dc. mOhm/m	1.825 MHz	3.5 MHz	7.0 MHz	10.1 MHz	14.2 MHz	18.1 MHz	21.2 MHz	29.0 MHz
36 0.19	587.1	169.23	104.70	64.63	51.47	41.75	36.09	32.87	27.31
34 0.23	400.6	127.14	80.28	50.59	40.66	33.23	28.86	26.36	22.01
32 0.27	290.7	100.94	64.82	41.51	33.58	27.59	24.04	22.00	18.43
30 0.31	220.5	83.30	54.24	35.17	28.59	23.59	20.60	18.88	15.85
28 0.38	154.8	65.72	43.51	28.61	23.38	19.37	16.96	15.56	13.10
26 0.46	104.7	51.11	34.39	22.91	18.81	15.64	13.72	12.61	10.64
24 0.56	70.1	39.92	27.24	18.34	15.12	12.61	11.08	10.19	8.62
22 0.71	43.3	30.02	20.76	14.11	11.68	9.77	8.60	7.92	6.70
20 0.91	26.2	22.55	15.76	10.80	8.96	7.51	6.62	6.10	5.17
18 1.22	14.7	16.42	11.58	7.98	6.64	5.58	4.92	4.54	3.85
16 1.63	8.3	12.05	8.55	5.92	4.94	4.15	3.67	3.38	2.88
14 2.03	5.3	9.52	6.78	4.71	3.93	3.31	2.92	2.70	2.29
12 2.64	3.1	7.24	5.17	3.60	3.01	2.53	2.24	2.07	1.76
10 3.25	2.1	5.84	4.18	2.92	2.44	2.05	1.82	1.68	1.43
5 5.38	0.8	3.48	2.50	1.75	1.46	1.23	1.09	1.01	0.86
1 7.62	0.4	2.45	1.76	1.23	1.03	0.87	0.77	0.71	0.61

Table 2 SWG round copper wire resistance

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"IK-STIC" by W2IK

(NOT TO BE CONFUSED WITH MY "IK-STIC 2" WHICH IS A MUCH BETTER DESIGN: CLICK HERE TO BE RE-DIRECTED)

The "IK-STIC" is essentially a multi-band vertical dipole antenna which is used in the field for quick set-up and quick band change. Since operation on each band requires antennas of different lengths, I devised a quick method of modifying the length to suit the band you choose to operate. There are no coils or traps. Each dipole band gives 100% radiation, thereby allowing maximum signal at maximum height.

==========

TO BUILD THIS ANTENNA YOU NEED:

1 - MFJ 1910 33 ft fiberglass mast (fig.1) the best choice

click above for link (telescopes down to 3.8 feet !)

OR

1 - WorldRadio SD-20 TELESCOPING MAST AND 1 PIECE of 12ft. 1 1/2 inch PVC pipe. (Cheaper than the MFJ purchase, just as good)

(insert SD-20 one foot into PVC and put a stop bolt in the PVC so the pole won't slide down further)

++++++++++++++++++++++++

You might wish to make an additional support (from PVC pipe) to hold the mast upright (or you may skip this and lash the mast to a tree or any NON-CONDUCTOR)

1 - either "CB" or 10 meter dipole antenna

(I bought a "CB" dipole, with center connector.. Workman Model BS-1)

(or you can build one using insulated wire) (see fig. 2)

1 - roll of STRANDED, insulated connection wire

(Radio Shack size #18-22 100')

Several Spade Connectors (Home Depot)

(both male and female with insulated covering)

BUILDING THE ANTENNA

Take the "cb" or ten meter dipole antenna, which already has a center UHF conector to accept a coax cable, and measure each side from the center. Trim EACH SIDE so that you have 8' 3 1/2" measured from the center. This makes a dipole of 16' 7" ..or a 10 meter dipole. At the ends of each leg, strip

Fig.2 Dipole with spade terminals

off about 1/4" of the insulation and attach a spade connector and crimp it AND solder it in place. see Figure 2.

FIG. 3

MAKING "BAND ADDERS"

Next, for each band other than 10 meters you wish to operate on, you will need to build "adders" for EACH leg of your ten meter dipole. For example: If you choose to make the dipole operate on 15 meters, the length of a 15 meter dipole is about 22' 2" or 11' 1" on each leg. Since you already have a dipole length of 16' 7", or 8' 3 1/2" on each side, you just need to make two "adders" of 2' 9 1/2". Cut two EQUAL lengths of stranded wire (each 2' 9 1/2") and add the opposite-type spade connector, so it mates with the ones you placed on the 10 meter dipole (see fig 3) crimp AND solder. You need only add it to ONE end of each "adder". When these wires are added to the 10 meter dipole, you now have a 15 meter dipole. Make other "band adders" using the same method. You see how easy it is to make "adders" for every band you choose to operate. I made them for 5 bands all the way to 40 meters.(see fig.4) DON'T try to skimp by trying to make "adders" on "adders" as it will make your set-up very complicated.

Adder sizes: Make TWO (one for each side) 15 meters: 2' 9 1/2" 20 meters: 8' 2" 40 meters: 24' 6" 17 meters: 4' 7 1/2" note: all approximate lengths.....mid bands

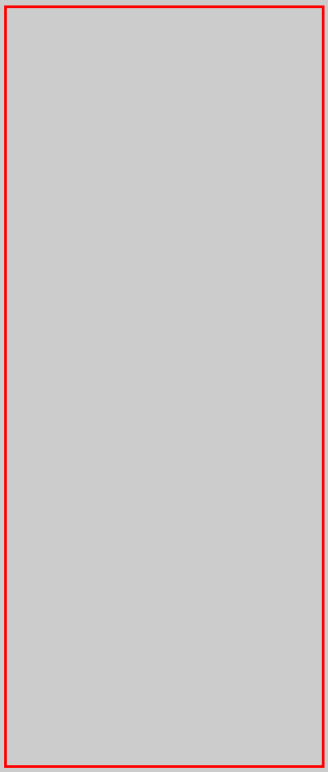
FIG. 4

SET UP AND OPERATION

Before you set your antenna up, mark which side of your 10 meter dipole antenna is connected to the CENTER of your coax connector. This side needs to be the highest in the air for best results. Decide which band you wish to operate on and connect your "adders" as needed (one on each side of the dipole). Take a rubber band and use it to hold the end of your "adder" to the TOP of the fully extended mast. Make it a very snug fit as this will hold the antenna as it runs down the mast.

You may wish to use a small amount of tape to hold the antenna to the mast just above the connector point to reduce any additional stress should you tug on the cable. Now add your coax to your antenna allowing it to flow down freely. The lowest end of your antenna wire should now be taped or rubber-banded to the low point on the mast. The antenna will not run to the ground unless you made a 30 or 40 meter antenna or used the Worldradio mast w/pvc. If you did, just tape where it meets the lowest point of the mast and let the additional wire rest on the ground in a straight line. (Be sure NO ONE comes in contact with the wire during operation.) Note: This is a vertical antenna. It can be used in tight spaces where an inverted "V" , etc cannot fit. (such as a limited area campsite) It needs no guys !

Any time you wish to change bands, just exchange the "adders" for the pair that matches your new band...I find this takes less than 2 minutes. So there you have the "IK-STIC" , all band antenna that weighs less than 5 pounds, needs NO tuner, and you can take it ANYWHERE ! I have worked hundreds of hams in dozens of countries while camping and at special events.



I hope you enjoy it too! Remember: It's called the "IK-STIC" by W2IK !

(the ideas described may not be reproduced without author credit)

++

VISIT MY "IK-STIC 2" WEBSITE TO SEE AN EVEN BETTER DESIGN. NO DANGLING WIRES AS THE RADIATING WIRE IS INSIDE THE TELESCOPING MAST. THE ENTIRE LENGTH OF THE ANTENNA IS 25 FEET, YET IT WEIGHS LESS THAN 5 POUNDS. CLICK HERE TO BE RE-DIRECTED!!!

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DESIGNED BY BOB HEJL - W2IK

(PUBLISHED IN AUGUST 2004 "WORLDRADIO")

The "IK-STIC 2" is a vertical, all band, antenna that is over 25 feet tall yet weighs under 5 pounds ! Using a tuner it can easily cover the amateur radio HF bands from 40 - 10 Meters. No unsightly wires as the radiating wire is inside the telescoping mast!

5992
HITOMETER

TO CONSTRUCT THIS ANTENNA YOU NEED:

ONE SD-20 Telescoping mast (WorldRadio sells these)
ONE 6 foot section of 1 1/2 inch PVC Pipe
50 feet of 20 or 22 gauge STRANDED, INSULATED Wire
ONE SO-239 Barrel Connector with washers and Nuts
ONE male and female push on connectors (see photos)
TWO Large (6 inch) Hose clamps (see photos)
Electrical Tape, Epoxy, Duct tape and asst. hardware.

ANTENNA CONSTRUCTION

FITTING THE INTERNAL ANTENNA WIRE INSIDE THE TELESCOPING MAST:

Take the SD-20 telescoping mast and remove the bottom cap by unscrewing it. Looking in you will see the sections nestled in place. Remove the rubber plug from the next to thinnest section so now all the sections are "open". Carefully take a 21 foot piece of 20 gauge, stranded, insulated wire and tie a very small knot at the end. Take the knotted end and insert it into the smallest section of the telescoping mast and using a straight wire made from a coat hanger, shove the stranded wire into the section as far as it can go. Then take a small amount of epoxy and glue the wire into place so it can't be removed from the top section. SLOWLY telescope out the entire mast, making sure that the wire slides inside easily. When the mast is fully extended you will have almost 20 feet of wire inside. Leave about 5 inches after the mast is fully extended and cut the wire. This will leave a 5 inch "play" to connect the wire at the bottom. Now CAREFULLY drill a small hole in the rubber base of the mast pointing out SIDEWAYS. Epoxy a push on connector into the hole. Solder another 4 inch piece of that same stranded wire onto the connector on the INSIDE. On the bottom cap of the mast, drill a hole that will allow you to half way insert, and tightly secure, that SO-239 barrel connector. Carefully epoxy it on the inside of the cap so it won't loosen. Next, solder the long wire that is in the mast onto the inner part of the SO-239 connector. Solder the wire from the push on terminal to the outer section of the SO-239 connector. Take the cap and give it about 7 COUNTER CLOCK WISE turns so the two wires are twisted. This way, when you screw the cap on, the wires will untwist in the mast. Tighten the end cap, but do not glue it.

WINDING THE PVC COIL SECTION:

Next take 25 feet of that same stranded wire and start to wrap it around the 1 1/2" PVC pipe at a point 14 inches from one end. (This becomes the top end.) MAKE SURE YOU LEAVE 8 inches of "free wire" before you start the coil wrap. Slowly wind the wire around the PVC pipe creating a coil, leaving a spacing of 1 1/2 - 2 inches from each turn. As you wind it down the pipe, you may wish to secure it every so often with electrical tape. The winding does not have to be exact, but keep it as evenly spaced as you can. One foot before the bottom, create a tight wrap of the wire, leaving no gaps on the turns. At the end, tape the wire to the PVC pipe. When you are done, wrap the entire coil in electrical tape so the coil stays in place. On the top end, solder a mating end of a push on connector so it can plug into the mast's side connector.

Wrap several turns of Duct Tape to the very top of the PVC mast. This will serve to offset the taper in the telescoping mast when it gets mounted to the PVC pipe. Using two adjustable hose clamps, carefully mount the very bottom of the telescoping mast to the top one foot of the PVC pipe. DO NOT OVER TIGHTEN. It takes very little compression to keep the mast in place. When you have done this, you can extend the mast out it's entire 20 foot length. To keep the entire antenna up-right, slip it over a 4 foot section of appropriate thin wall steel tubing that has been pounded in the ground about one foot. The lower coiled section of the antenna on the PVC pipe will then be slightly "ground coupled". This helps with the antenna's operation on 30 and 40 meters. Plug in the lower coil (The PVC pipe) into the male connector on the side of the telescoping mast. The SO-239 connector is where you screw in your coax cable to your radio. Make a few windings of whatever coax you are using at the connector point and tape them tightly together to prevent RF from returning on the coax shield. Connect the other end of the cable to your tuner and you are all set to go !!.

To dis-assemble the antenna, just remove the coax, loosen the hose clamps and take down the mast after unplugging the PVC coil plug. CAREFULLY retract the mast and the internal wire should slowly coil down into the masting. DO NOT FORCE THE SECTIONS. A few gentle jiggles and a twist or two will do the trick. After several uses it will be easier to retract the sections as the internal wire will have "memorized" how to coil up. You can even store the telescoping mast in the PVC pipe by making a small slot at the bottom of the PVC tube to accommodate the connector that is on the side of the telescoping mast . The antenna is very simple, light and works well when tuned properly. My first contact was on 15 meters when I spoke to Siberia. I have used it on all the bands it covers and have also made an adapter so it mounts on the ball hitch of my truck. This is great when you are parked and can't make a hole in the ground. (NOTE: If you wish to make an "IK-STIC 2" that covers 160-10 meters with a tuner, use a 7 FOOT PVC PIPE instead of the 6 ft. pvc and coil 35 feet of wire around it using 1 inch spacing between wraps and two feet near the end increase the spacing until you run out of the wire and the end of the coil wrap is four - six inches from the bottom of the PVC pipe. Any longer coil winding that this will make it difficult to tune the antenna on 10 meters.) (Use the rest of the antenna building dimensions as outlined above.)

REMEMBER.... it's called the "IK-STIC 2"

Designed by Bob Hejl - W2IK

This antenna has been used at Field Day operations, Special Events Stations, JOTA Events and County Activations with great results.

IK-STIC2 STORED IN IT'S LOWER COIL SECTION

LISTEN FOR MY W2IK/B BEACON STATION OPERATING ON 28.242 MHZ. CLICK HERE TO SEE PICTURES AND INFO ON MY BEACON STATION

page created with Easy Designer

RADIO HABANA CUBA

THE EVER-POPULAR INVERTED "L" ANTENNA

BY ARNIE CORO (CO2KK)
Host of "Dxers Unlimited"

SEND YOUR COMMENTS, QUESTIONS AND IDEAS DIRECTLY TO ME AT:
arnie@radiohc.org

Ever wanted to install a shortwave receiving antenna FAST?!!! Then I am almost sure that you will want to learn more about the "Inverted L" antenna... About the easiest effective skywire to build ...

Just two supports... may be two masts, a mast and a tree, a mast and a nearby building... it need not be perfectly horizontal above ground... as a matter of fact if the inverted "L" is installed in a slightly tilted angle it seems to work better.

Although strictly speaking a true inverted L has the downlead connected to one end of the horizontal section... my version calls for connecting the downlead (a single wire) to a point about 20 percent from one end.

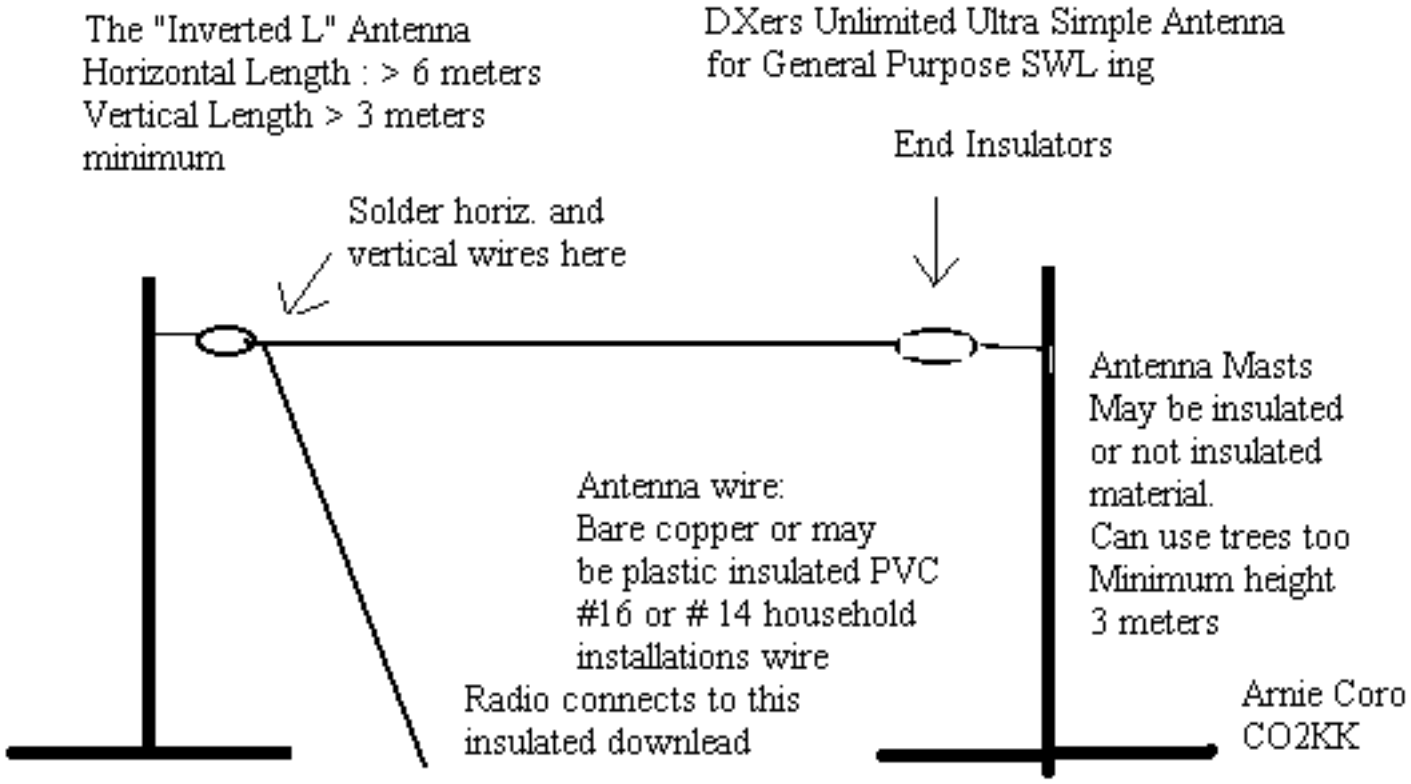
The antenna should be connected to your receiver via an antenna tuner with a real wide matching margin. For a 10 to 15 meter horizontal section, strung between two masts of about 5 to 7 meters each, you may expect very nice performance on frequencies extending from about 3 to 30 megahertz.

Using the inverted "L" for transmitting, does require the installation of a good ground system or a "floating counterpoise", a device coming from the early days of radio that provides an amazing improvement in performance of many antennas... More about the "floating counterpoise" soon right here at this WWW site...

You may want to build your inverted "L" with copper wire of no less than no. 16 gauge.. I prefer to use PVC covered copper wire of no 16 or no 14, same type as used by electricians for home installations.

Overall antenna length is not really critical, but the inverted "L" does need have a

horizontal section of no less than 5 to 7 meters and a vertical section of no less than 3 to 5 meters to work effectively. The most tested version here has a spacing between masts of 20 meters, a height above average terrain of no less than 5 meters and I used a single wire floating counterpoise, installed below the horizontal section of the antenna.



Do remember than in order to obtain optimum performance from this very simple, yet effective, short wave antenna, you do need to install it properly, solder the antenna downlead to the horizontal section, and use an ANTENNA TUNER...

Wish you success with this rather simple antenna project...

Doubts??? Then send a message to arnie@radiohc.org

Havana, Cuba
25 December, 1997

Send e-mail to Arnie Coro

Back to RHC Homepage

The KGØZP Super Linear-Loaded Inverted V

How do you fit a full length 160 meter antenna into a 40 foot deep yard?
Install the KGØZP Super Linear-Loaded Inverted V, of course!

The Super Linear-Loaded Inverted V allows you to select your two favorite and most used frequencies as band centers, and offers excellent bandwidth as well.

The designers installation centerpoints are set at 1.840 the center of the DX window or calling frequency and 1.875 at an SWR of 1.1-1
The bandspread at 1.5-1 is 20 kc at each center or
40 kc to an SWR of 2-1

The linear-loading section is mounted near enough to the ground that you can easily install or remove precut pigtails using an alligator clip to move the center point temporarily to make that sked, or you can tune the whole 160 meter band with a simple antenna tuner.

The formula for construction is the same as for any dipole 468/fMhz and the design will work equally well on 80 or 40 meters.

Reader Please Note: The photos on this page are greatly compressed to insure quick downloading.
The link accompaning each photo will cause the uncompressed full size photo to download.

It would be impossible to show an overall view of the whole antenna system in such a way that you could see the wires.

With that in mind, I have taken a couple of daylight pictures as well as flash night pictures which allowed me to highlight the wires against a black sky.

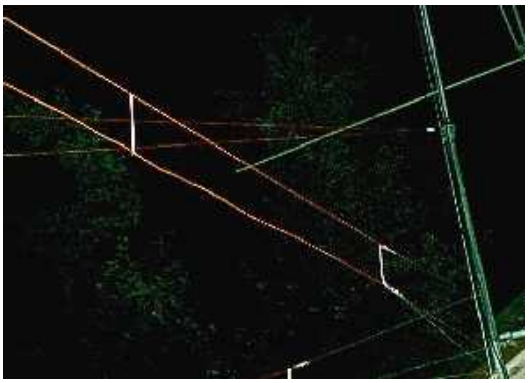
Lets take a thumbnail look at a daylight view of the Rohn 50 foot push-up pole utilized for this installation.

This is quite a busy photograph, so I will name the items in descending order to help pinpoint the areas relavent to the Super Linear-Loaded Inverted V. Link to view photo in hi-res!

The white stick at the top extending off the upper edge of the photo is a Diamond Tri-bander, model X3200A, fed with 9913 coax.
Directly below the tribander is the first guying ring. This is a dummy ring is used to ground and hold the upper wire of a Wermager type Broadband Sloper. In the full sized photo you can just barely see this wire extending left forward.
The highest of the three actual guying rings (second one down from the top) is at the 40 foot high point on the tower, a pulley is connected to this guying ring and through it is threaded a high quality flagpole rope. This rope can be used to raise and lower the balun for experimenting with different antenna designs, without changing the position of the push-pole.
The Super Linear-Loaded Inverted V, the topic of discussion is connected to this balun. The tension on the wires is so great that it holds the balun outward, the line descending from the bottom of the balun is Super Mini-8 coax. The pull of the Inverted V's two antenna wires is offset by two guywires on the other side of the guying ring. I have a wide yard and could keep the apex slightly wider than 90 degrees.
The next guy ring down is at the 30 foot height point and has nothing more three guy wires.
The lowest guy ring visible in the photograph is at the 20 foot level, and is the feedpoint for the Fan portion of a Wermager type Broadband Sloper originally used on 160 but now cut for 80 meters. I would like to mention again that the upper wire of the Wermager type Broadband Sloper is the wire grounded to the push-pole at the dummy guy ring at the very top. The other end of this wire is tied to the same location as the guy wire coming from the 30 foot guy ring, which places the metal guy wire directly between the driven Fan and the grounded sloper element of the Wermager type Broadband Dipole.

This photo is a view showing the East outbound leg of the Super Linear-Loaded Inverted V. In the upper left of the photo, you can see the single sloping wire descending from the balun. The two horizontal wires below comprise part of the Linear-Loading section. Link to view photo in hi-res!

The remaining photo's were taken at night using a flash, then enhanced to make the wires more prominent.



In this first photo, you can easily see that the Linear-Loading section is a good distance away from the push-pole. I used a distance of 5 feet in this installation, but you can reduce this distance to 3 feet if necessary. The vertical spacers are drinking straws, one nested inside the other for greater strength, a nylon tie line is fed through the center of the straws. Link to view photo in hi-res!
The upper tie off lines are located 1 foot above the standoff mounted to the eave. The lower tie off lines should also be located at or just above the standoff. Original twine used during initial setup is still visible and should have been removed before taking the photo.

The white insulator some 5 feet above the linear loading section is the feedpoint for the Fan portion of the Wermager type Broadband Sloper.

SPECIAL NOTATION: The upper wire and the lower wire of the linear-loading section are connected at this end only. This will be described in greater detail in the construction notes section. Link to view photo in hi-res!

These last two photos show the East and West outbound ends of the Super Linear-Loaded Inverted V.

If you look closely, you can see the pigtails wrapped around the lower nylon guys.

In the photo on the left of the East outbound end, the wire from the balun down to the insulator is not visible. The upper guy wire from the insulator is metal, the lower is nylon. An additional twisted wire above the insulator is used to lift the linear loading section slightly higher than the existing guy mount allows, I did this only because I had something above each outbound end to tie to. Link to view left photo in hi-res!

The right photo shows slightly greater detail and the antenna wire from the balun is clearly visible. There are no electrical connections at the outbound end of the linear-loading section. Link to view right photo in hi-res!
Again, if you look closely, you can see the pigtail extension wrapped around the lower nylon guy.

Construction Notes:

The construction of the Super Linear-Loaded Inverted V is quite simple and follows the same formula as for a standard dipole. 468/fMhz (FourHundredSixtyEight divided by the Frequency in Megahertz give you the total length in feet for a 1/2 wave dipole). In practice you will use two wires, each being 130 feet long. The antenna shown in the photos is made from 14 guage insulated copper wire. I ran out of insulators and am temporarily using loops made from 1/4 inch nylon cord at the push-pole end of the linear-loading section. If you start with two 130 foot long wires, you do not have to measure the turning or ending points, they will fall in place depending upon the height you install the horizontal portion of the antenna.

The height of the feedpoint can vary considerably, however, the higher the better. If you can only go up 30 or 35 feet, the antenna will still perform almost as well, but you will require a further horizontal distance to work in. With the feedpoint at 40 feet, the sloping elements are roughly 60 feet long at 40 feet away from the push-up pole. Any distance from 40 feet to 80 feet in length seems to make little difference in the performance of this antenna.

A balun is not an absolute necessity, however, if space constraints keep you from obtaining a spread between the sloping wires greater than 90 degrees, then a balun is strongly suggested.

For all practical purposes, installation is the same as for an inverted V, a feed type insulator or balun is assembled with the coax and both of the 130 foot long wires. Connection of the coax is the same as for a standard dipole. The center conductor goes to one wire and the shield to the other wire. The feed insulator or balun is now hoisted to it's permanent position on your push-pole, tower, gable or wherever you are mounting your antenna.

I will describe the finished setup first, then give some quick install tips immediately following.

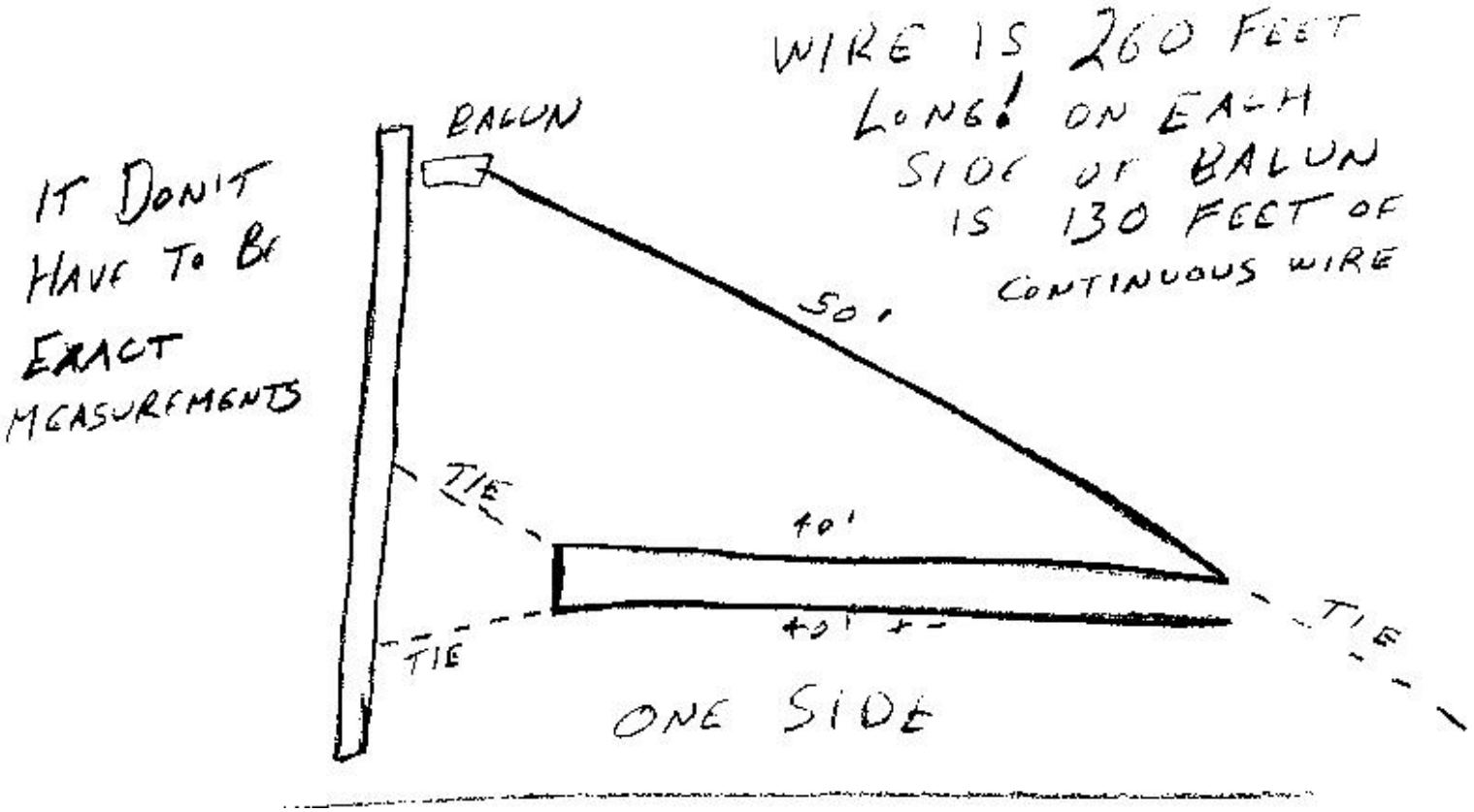
One side of the antenna is assembled first and will require a minimum of three insulators or you can use nylon mounts by giving the wire a twist to form a loop to connect the nylon rope to.
The antenna wire is brought out across the yard and the end of the wire is passed through an insulator at the outbound end of the antenna. The antenna wire is then brought back toward the push-pole and passed through the second insulator. The second insulator should be installed about 6 feet away from the push-pole temporarily. The wire then passes through a third insulator 6 to 8 inches below the second insulator, and then proceeds back out to the outbound end.

The ideal installation would have the feedpoint 50 feet above ground level and the linear-loading section 20 feet above ground level. In practice, the feedpoint can be as low as 30 feet above ground level and the linear-loading section as low as 10 feet from the ground without affecting performance to drastically. In any case, the linear-loading section should be maintained as close to horizontal as possible.

The idea is to determine at what height you wish to install the horizontal linear-loading section, and then adjust the insulators and distance between the push-pole and inside end of the linear-loading section so that the first insulator and end of the wire are on the same plane. And you'll really love this, give or take 2 or 3 feet on that tail, preferably give a few feet extra for trimming to resonance.

Any non-metal spacer can be used to maintain distance between the elements of the linear-loading section! No need to go overboard here, it's close enough to the ground you can replace the simple item I use in Quick Tips below.

Simple Sketch showing path of antenna wire on one side of balun.



Quick Tips:

Here is how to install this antenna system and have it up and running in well under an hour, exluding push-pole installation of course.

1. Install the wires to your balun and connect the coax, hoist it into position and secure firmly.
2. Pull one wire out taught to your guy mount and secure temporarily.
3. Take the loose end of this same wire and secure it to the push-pole temporarily at the desired final installation height. Note: This end of the wire will be at the other end of the yard when finished.
4. Go back to the outbound end of the antenna, using a stepladder if necessary, pull the wire so that it is taught from the push-pole and lift the wire until it is perfectly horizontal.
5. With a magic marker or piece of tape, mark the sloping wire at the point where the wire from the push-pole crosses the sloping wire. This is where the outbound insulator is to be installed.
6. Install the outbound insulator and tie off (guy) this portion of the antenna to permanent tension.
7. Disconnect the loose end of the wire from the push-pole, pass it through an insulator and bring the loose end out to the outbound end and temporarily connect it to the now secure insulator outbound insulator, leaving 2 to 3 feet of loose pigtail extending through this insulator.
8. Pull the wires from the insulator toward the push-pole until they are in balance at equal lengths from the outbound end, at this bend take the insulator and give it a twist, release the loose pigtail at the outbound end, then secure the push-pole end of the top linear-loading wire and it's insulator to the push-pole at it's permanent position.
9. Install your first spacer (see spacers below) by passing the loose end of the wire through it, bring it up snug to the insulator, install the second insulator and give it a twist to hold it secure. Tie this insulator off to the push-pole as well, trying to keep the spacer vertical.
10. Install additional spacers, one about every five feet, to maintain separation of the upper and lower linear-loading sections.
11. When you reach the outbound end, install your last spacer at the insulator, form a loop in the lower wire for attaching a nylon tie off rope. Pull the rope to tighten the lower linear-loading wire and tie off to the guy mount. Wrap the remaining 2 or 3 feet of pigtail around the nylon rope.
12. Duplicate the above instructions for the other leg of the antenna!

SPACERS:
I take simple drinking straws to use for spacers. But reinforce them to two thicknesses for all the spacers except the inbound spacer near the push pole, this one is three or four thicknesses.
To strengthen a soda straw, you merely slit another soda straw end to end and insert it inside of the whole unslit soda straw.
You only need to pass the unslit whole soda straw over the wire for the inbound spacer, the other two or three slit straws can be added right over the wire and then slid into the whole unslit straw.

Joining the straws to the wires:
I take 24 inch long pieces of braided nylon, like mini-blind sash cord, fold it in half, hold it on one side of the top wire with the loop up and open, then I pass the loose ends of the rope over the wire and through the loop, pull it snug and let the ropes hang until I get all the ties (ropes) installed. By taking a 12 inch long piece of 18 or 20 guage wire and bending a sharp turn at one end, you can use that to slip the ends of the ropes into and slide the soda straw over the wire pulling the rope down through the straw. Place the loose ends of the rope one on each side of the lower wire and tie a square knot, drawing the wire up tight against the soda straws. Don't worry if they are atilt right now, you can true them up later.
After the antenna is assembled, tied down nice and snug, and tuned. I will adjust the soda straws to vertical and place a dab of something like plasti-dip-your-grip or dumb gum over the tops of the straws to hid the nylon braid from UV and weather. I usually don't worry about the knot end or lower end of the straw.

Tune-Up:
If I have obstructions in the yard that interfere more with one leg than the other, I will make the shield side of the coax go to that side and the center conductor to the clearer of the legs. Now I tell you!
I normally tune both legs of the antenna first to 1.840 MHz, then I will continue to cut the clearer leg of the antenna all the way down to 1.950 MHz. I will then make a pigtail with an alligator clip on it and retune that leg of the antenna to 1.875 MHz. You can wrap the pigtail around the nylon guy and slide it down at least 6 inches away from the tuned pigtail end and it won't interfere with your upper band setting. In my case, I leave the pigtail connected as 1.875 MHz is one of my popular areas of the band for ragchewing. I may be over 50, but to really fit in up at 1.950, I need another 40 years under my belt, Hi Hi..... You may make other pigtails to center in other areas of the band also. It works on either leg of the antenna!

TTUL - 73+ de Gary - KGØZP

Questions or comments may be e-mailed directly from the home (index) page.

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Document Revision Date 11/17/99 Applicable To This Page Only

Dipole and Inverted V Antenna Calculator

Enter The Frequency For The Dipole/Vee Antenna Calculation
Use entries like 7.200, 7.2, 144.200, 144.225, etc...

Divided by Freq MHz

Percent smaller for the Inverted Vee

Your dipole's total length is feet

Each leg of the dipole is feet

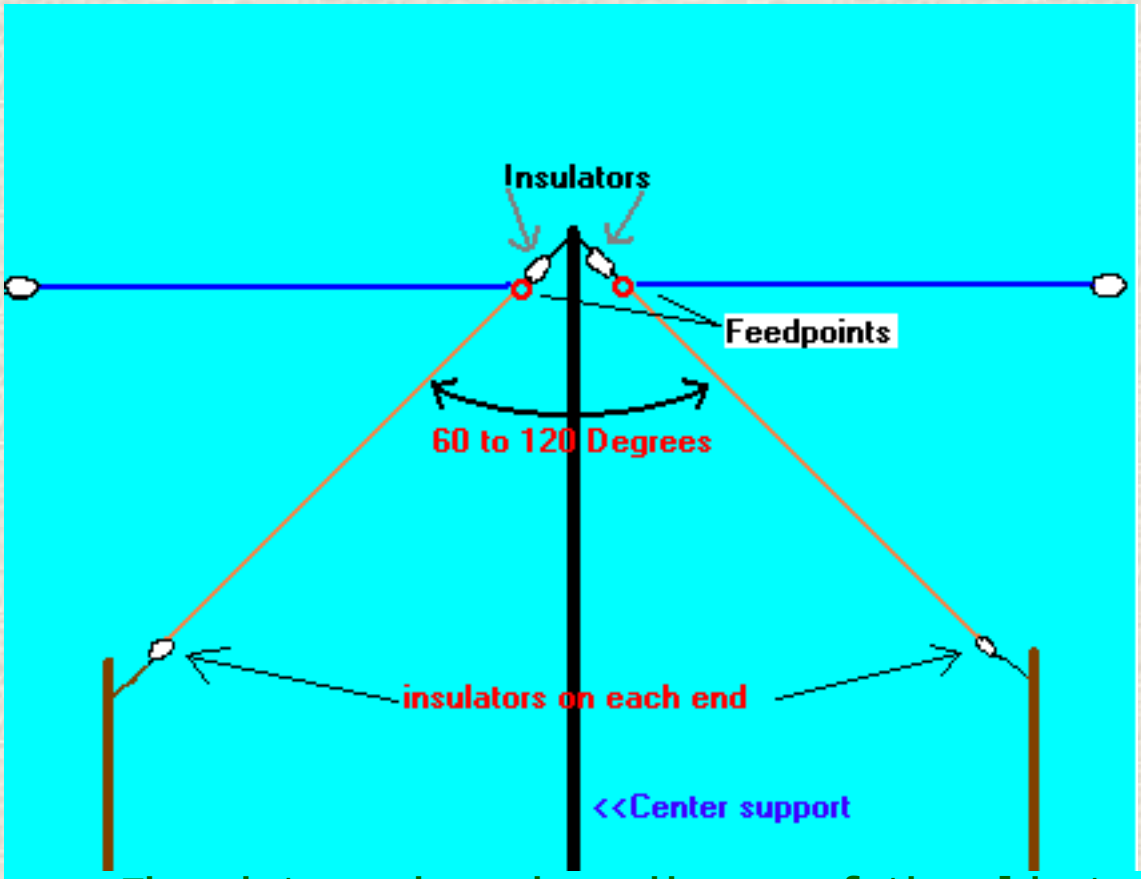
Your Inverted Vee's total length is feet

Each leg of the Inverted Vee is feet

HAM UNIVERSE

THE DIPOLE AND INVERTED VEE CALCULATOR WITH DRAWING
TWO CLASSIC ANTENNAS

BACK TO ANTENNA LAB



The picture above is a diagram of the old standby CLASSIC antennas....
the Dipole and the Inverted Vee.
They are shown superimposed on each other in the drawing.
The Blue line is the example of a dipole.
The copper colored line is the inverted vee configuration.
The inverted vee is is just a lazy dipole that can't hold it's arms up and is about 4 to 5% shorter on each half!
Both antennas can be constructed just about anywhere using any type of wire you may have and can be used with 50-75 ohm coax or open wire ladder line and a tuner. The use of an rf choke balun is recommended!
You could use the calculator to build a 75 meter dipole and then calculate an inverted vee for other bands suspended under it from the same support and feed both with the same line to xmtr. A little experimentation may be required for adjustment of the SRW, tuning the 75 meter dipole first for lowest SWR and then the inverted Vee.
EXPERIMENT! EXPERIMENT! EXPERIMENT!
Learn more about the inverted vee compared to the dipole
click here to leave this site to learn more!.

Back to Antenna Lab

Antennas & More for HF antennas

Confused about length conversions? Download this neet length converter calculator. (It has been checked to be virus free!) Comes from a government website! Also see chart below

Length Conversion Factors

To convert from	to	multiply by
mile (US Statute)	kilometer (km)	1.609347
inch (in)	millimeter (mm)	25.4 *
inch (in)	centimeter (cm)	2.54 *
inch (in)	meter (m)	0.0254 *
foot (ft)	meter (m)	0.3048 *
yard (yd)	meter (m)	0.9144 *

More conversion Tables, Charts, Calculators etc

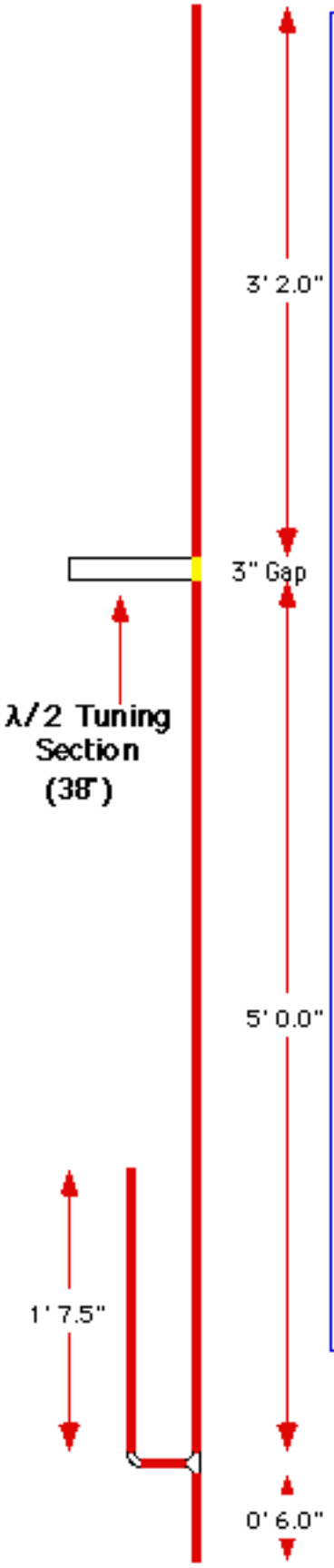
Metric and Imperial Length Conversion Calculators Charts Tables More

The Super J-Pole Antenna(Collinear Design)

Below is a Javascript for calculating the lengths of tubing to be used for the construction of a Super J-Pole. The antenna I constructed was made of 1/2" tubing.

Frequency

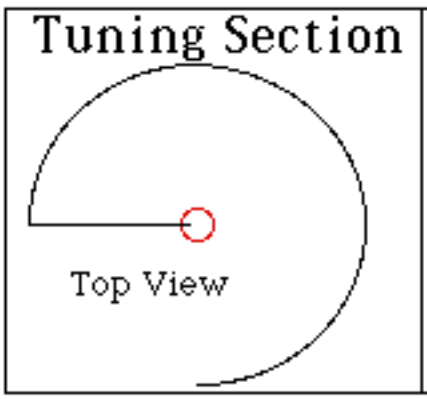
Bottom Section	inches
Top Section	inches
Matching section	inches

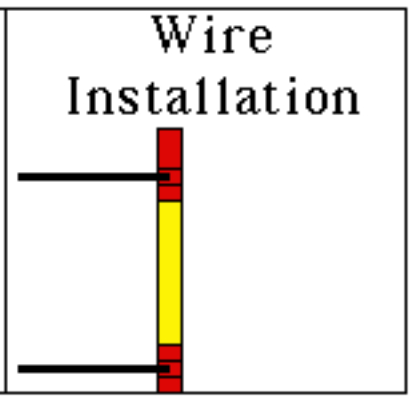


Super 2 Meter J Pole Antenna (a.k.a. Collinear Antenna) with helpful modifications by Aaron Schmitz, KB0YKI

Construction Details:

1. Cut copper to proper lengths.
2. Flux and solder bottom J Pole section.
3. Cut a 38" piece of 12 gauge wire.
4. Solder a 1/2" coupler on the top J pole section and bottom of the top λ/2 length of pipe.
5. Drill a 1/8" hole into each coupler through the pipe.
6. Strip back 3/8" of insulation on wire.
7. Insert wire ends into holes and solder.
8. Bend wire into a 3/4 circle around pipe with a 6" radius.
9. Insert 1/2" X 4" dowel rod into coupler.
10. Completely seal all areas from moisture. Paint it if you would like.
11. Attach coax and have fun!!





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The End-Fed Halfwave Antenna

According to AA5TB

The End-Fed Halfwave Antenna is nothing more then a resonant dipole antenna that is fed at the high voltage point (high impedance end) instead of in the center, which would be high current point (low impedance). This antenna is as old as radio and is probably best known as the Zepp Antenna named after the airship is was commonly used on, Zeppelin. Since in the early days of radio the wavelengths used were very long the antennas were also long. An end-fed antenna that could be trailed behind the airship was much easier to implement then a center-fed dipole or a 1/4 wavelength antenna that required and extensive ground plane to function properly. Amateur radio operators had found that when an antenna was cut 1/2 wavelength on the lowest frequency of operation it could still be end fed at a voltage loop at all harmonic related frequencies. And since all of the amateur radio frequencies at one time were all harmonically related this made an ideal all-band antenna.

The main reason I like the End-Fed Halfwave Antenna is that it makes an excellent portable antenna for the field. I often take my my QRP gear out to the field during camping trips and the less antenna gear I need to bring the better. With the End-Fed Halfwave Antenna all I need is a short (or no) feed line to my coupler, a half wavelength of wire, and a bit of rope and I'm all set. The feedpoint usually ends up near the ground and if I can I prefer to run the wire vertically although I've had very good success with it run out horizontally only about 10' up.

When used as a vertical the End-Fed Halfwave Antenna is ideal because like the dipole antenna, it is self-completing (because it IS a dipole!). That is, it doesn't require a return path or "counterpoise" to properly radiate. This makes it an ideal portable antenna. There are many portable antenna designs in use that really require an extensive ground plane to be efficient, such as the 1/4 monopole or loaded versions of such. Yes, even though the 1/4 monopoles with a few radials will work and often work good, there is no escaping the fact that power is being wasted in the ground and I don't that feel carrying around a bunch of radials is all that much fun.

Now the last statement about this antenna requiring no radials has been debated extensively. In ON4UN's book, "ON4UN's Low-Band DXing" he has modeled the vertical End-Fed Halfwave Antenna and has shown that a radial system with numerous, very long radials will increase the performance of the antenna. He indicates how currents are indeed setup within the ground and losses will occur if radials aren't used. Therefore, he concludes that radials must still be used. I agree with his data but I disagree with his

conclusion. I disagree because the currents that are setup in the ground are maximum at about 1/4 to 1/3 wavelength from the base of the antenna, unlike the 1/4 monopole that has a maximum ground current density right at the base. The other fact is that with the 1/4 wave antenna, this return current in the ground is required for the antenna to function. For the 1/2 wave antenna, the ground current is there simply due to the ground being in the near field of the antenna. The higher you raise the antenna, the less current there will be in the ground. In practice I have found that the vertical End-Fed Halfwave Antenna with no radials outperforms a horizontal dipole at 20' on 20m on short and long paths. Sure, many radials one wavelength long may improve the performance a extra fraction of a dB but if it already works great why go through the trouble during portable operation? Now if you were creating a contest station on your ranch you may want to provide this elaborate ground system.

Others have suggested that one 1/4 wavelength radial should be used to keep the coupler at a low impedance. This might make you feel better about keeping RF at bay near the coupler but in practice I haven't needed it. In fact, if this does make a difference then current must be flowing in this 1/4 wavelength of wire and since there is only one wire it will radiate. I have found that this is usually caused by the antenna not really being an electrical halfwave in length. Having a radiating part of your antenna laying on the lossy ground can't be all that great.

Now there are several ways to feed a End-Fed Halfwave Antenna. These methods can be put into two categories:

1. Tuned circuits using lumped components.
2. Transmission line transformers.

The idea is to convert the very high impedance of the antenna (usually around 5k ohms in real life) down to the relatively low impedance (50 ohms) of the transmission line. Now it should be known that if the antenna was infinitely thin, in outer space, and was exactly resonant no energy could be coupled into this antenna using this method (this is what the physicists will tell you). If things were perfect you would have to couple into the antenna just slightly in from the very end of the antenna. In other words, move the feed point from the very end to a foot or two from the end. This would mean that the antenna would be on one side and a very short return wire would be on the other side. This would present a small amount of capacitive coupling from one side of the circuit to the other and RF current would flow. "Luckily", things are never perfect and this situation almost always exists anyway. Therefore, a small return wire may be necessary although there is usually enough stray capacitance from one side of the circuit to the other to satisfy this requirement without any additional return wires.

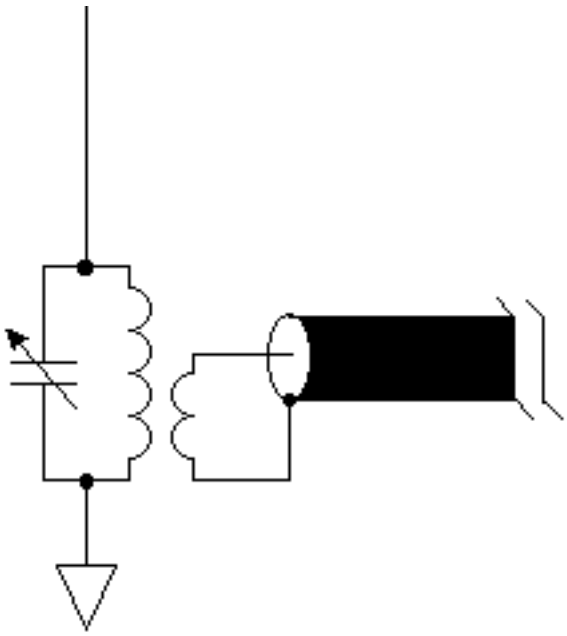
When using lumped components a parallel circuit is connected to the end of the antenna and tuned to resonance at the operating frequency. This parallel tuned (tank) circuit also

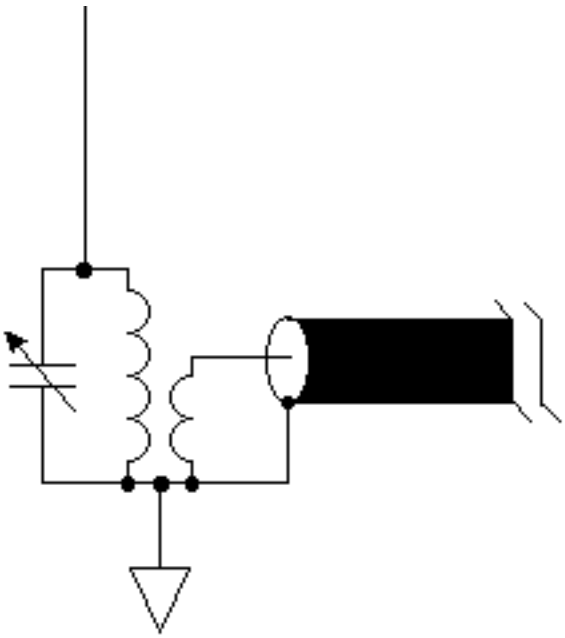
presents a high impedance and easily couples to the antenna. To step the impedance down to 50 ohms you can use either a small winding around the "cold" end of the main coil in the parallel circuit to use as a primary winding or simply use a tap on the main winding just up from the "cold" end of the main coil. The latter approach is essentially an autotransformer and the tap that presents the lowest VSWR at the resonant frequency is the correct one. A Cushcraft Ringo is another example of this type of antenna. It is essentially using the autotransformer parallel tuned circuit approach but since the frequency is so high the main coil is just one turn of aluminum.

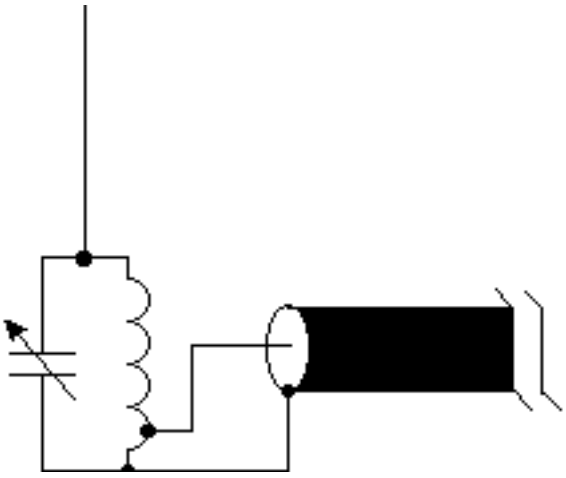
When using a transmission line as a transformer a length of line that is roughly 1/4 electrical wave in length is used. The velocity factor of the line must be taken into account and can vary from one brand or type of transmission line to another. Essentially what we do is use a section of transmission line to transform the high impedance of the antenna (~5000 ohms) down to vicinity of 50 ohms. Then, across the feedpoint we place a short section of shorted transmission (stub) line to finish the transformation. We now have what is best known as a J-Pole antenna. It is often described as a shorted section of a 1/4 wavelength transmission line that is tapped at the 50 ohm point but Gary E. O'Neil, N3GO has done extensive studies regarding this type of feed and his text on the subject is recommended reading. We usually would want some way to ensure the external currents are not coupled onto the coax with some form of a balun although most designs do not incorporate one. J-Pole antennas are very common for VHF/UHF operation although most users don't realize that they are using an End-Fed Halfwave Antenna. I have used J-Pole antennas extensively but I believe Gary can describe them much better then I can.

Here are various methods of feeding an End-Fed Halfwave Antenna

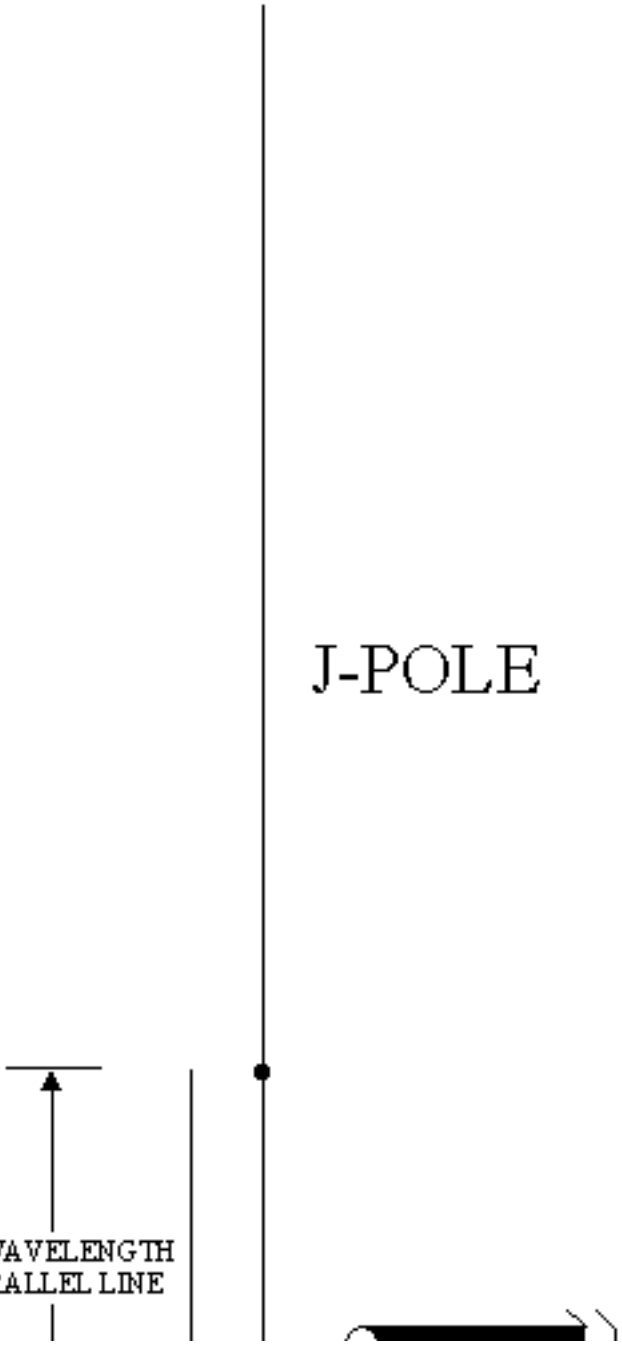
(The return connection can be ground or just enough conductor to present a bit of stray capacitance)

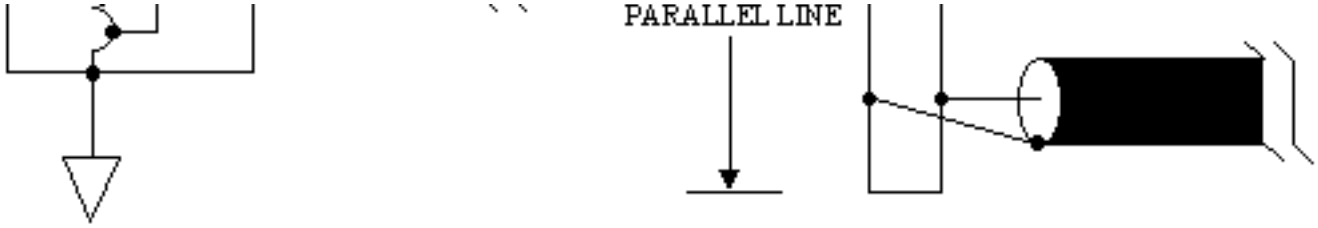


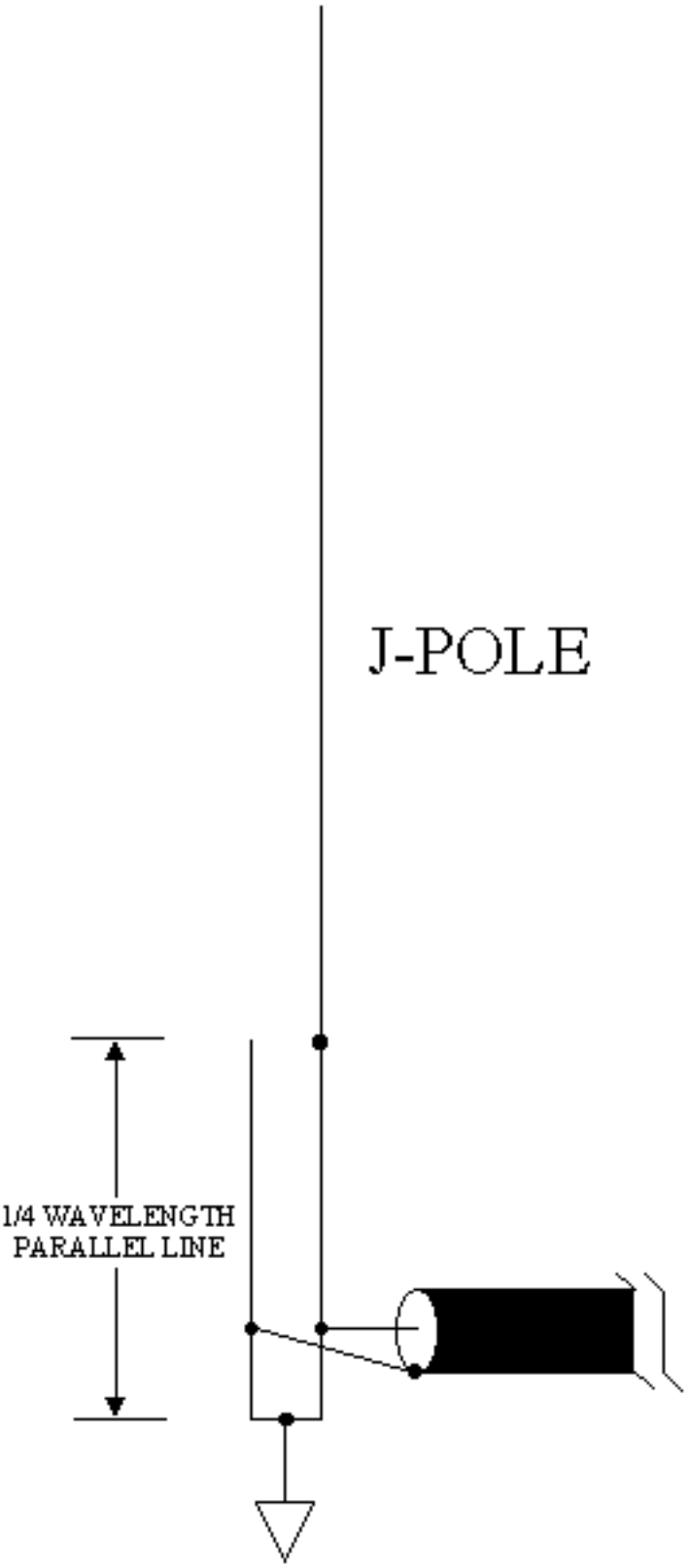




J-POLE




PARALLEL LINE


J-POLE
1/4 WAVELENGTH
PARALLEL LINE

EMPIRICAL RESULTS

During the Lockheed Martin Recreational Association Amateur Radio Club's (whew!) year 2000 Field Day we operated 2A-Battery QRP. We had several wire antennas at our disposal but I would like to give you the EMPIRICAL results of an antenna test that I squeezed in.

On 20m I put up the following two resonant antennas with the following conditions:

1. End-fed vertical halfwave dipole with the bottom up 3' from the ground. A link coupled parallel coil/capacitor circuit for impedance matching to coax. NO RADIALS or GROUND SCREEN.

2. Center-fed (via choke balun) horizontal halfwave dipole up 20' broadside north/south.

3. Same length of RG-58 feedline with losses measured at 1.1dB each.

4. The horizontal dipole was clear of support branches and foliage by several feet.

5. The vertical dipole was clear except that the top couple of feet were thread into the foliage.

6. The antennas were CROSS polarized and separated by at least one halfwave length so mutual coupling should have been minimal.

7. Soil was probably average-good. The ground is normally solid limestone just below the surface but this area probably has top soil sediment from the West Fork of the Trinity River judging by the giant pecan and cedar elm trees.

Now, I have used a horizontal dipole at similiar heights for many years at home and it has always been my a very sure performer, stateside and DX. I didn't expect any better performance without going to a more directional gain antenna or an increase in height. I have used the end-fed halfwave style of antenna for years during my portable excursions and it has always performed well and I have never used radials with it. My expectations were that the vertical dipole would perform better on stations greater then about 2000 miles and that the dipole would perform better on the shorter paths, especially since the peak of its major lobe would be straight up at that height. I was incorrect.

The end-fed halfwave (vertical dipole) routinely outperformed the horizontal dipole on <2000 miles paths by AT LEAST an estimated 6dB except during very short periods of time when the signal would be predominately horizontally polarized. Mysteriously, there

was not as dramatic of a difference on DX signals although the vertical was almost always better.

The dB delta was estimated by S-meter readings on a rig that I had measured to deflect about 1 S-unit every 4dB. The vertical dipole was often better by about 2 S-units so 6dB is probably a conservative figure.

Sprodaic-E propagation was very prevailent during that Field Day and on 20m I was working stations at times only 150-200 miles away on skywave yet the end-fed halfwave was still outperforming the horizontal dipole.

I kept looking for something wrong with the horizontal dipole but could find nothing. What could go wrong with a dipole anyway :-) ?

I admit this wasn't a purely scientific test but I believe it was a reasonable A/B test.

Bottom line...

1. A vertical end-fed halfwave dipole outperformed the tried-and-true horizontal dipole at my Field Day site even though I would not have originally suspected that this would be the case.

2. Radials and a ground screen are NOT required but you may use them to gain that extra fraction of a dB (L.B. Cebik, W4RNL) (QST, July 2000, pg38) if it makes you feel better. There is little current in a ground screen for this antenna until about 1/4 wavelength from the antenna axis, so make your ground screen BIG and carry LOTS of wire. A ground screen may significantly detune a halfwave antenna from resonance since you will be placing a large conductive mass near the high impedance end of the antenna. You may create your own problems. If the antenna is not near a halfwave length, more current will flow near the base of the antenna requiring a better ground screen.

3. I wish I could put one up at my house vertically.

I would like to hear from others who can perform this type of A/B test using skywave signals for the vertical end-fed halfwave dipole versus other field ready antennas. I really did not expect this kind of difference between the horizontal dipole and the vertical end-fed halfwave dipole antenna during portable use and I still would like to have more measurement data.

By the way, we only made about 330 Q's on CW (15/20/40) and 240 Q's on SSB (10/15/20/40) with our solar charged battery operated QRP gear. We could have made more but I spent a lot of CW operating time Elmering some new and potentially new hams about ham radio and this was worth more "points" then the QSO's. I did have one short

rag chew with a station in Tokyo, Japan using the end-fed vertical halfwave.

Related Links

Here are some other ideas on the same type of antenna:

- From a J to a Zepp - The truth and its consequences - Gary E. O'Neil, Raleigh, N.C. (N3GO)
- An End Fed Half Wave Antenna, by the late G3YCC
- END FED ANTENNA IDEAS FOR FIXED OR PORTABLE,by G3WQW
- J-POLE ANTENNAS - by Mike Walkington, VK1KCK
- Portable End-Fed Halfwave Antenna - by OE3MZC

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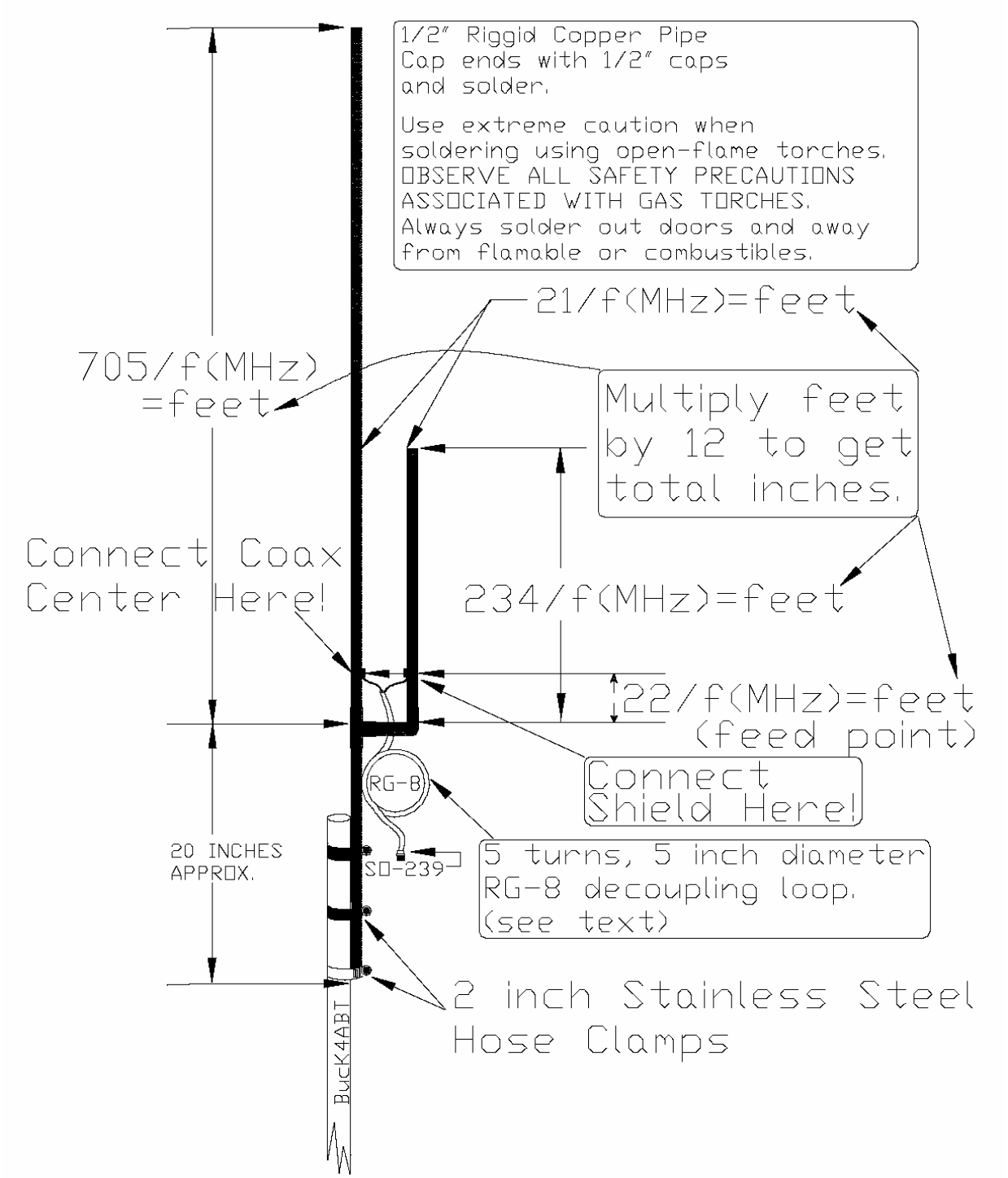
aa5tb@yahoo.com

BACK HOME

Last updated on May 21, 2004.

On line since July, 1998



Notice how the decoupling loop is formed from a length of the coax coiled into a 5 turn, 5 inch diameter loop. The coil is then attached it to the mast as shown above. For the final touch, attach an SO-239 (female) coax connector; see text.

A more efficient decoupling transformer is available from:

BUX CommCo, at: www.BUXcommCo.com

AN OUNCE OF LIGHTNING PROTECTION:

The shorted matching stub of the J-pole also appears as if it is a short circuit across the antenna terminals of the connected radio or transceiver. There is also and ounce of protection from static discharges via the antenna port even if you don't make a practice of disconnecting your antenna before a storm. To make life easier and, should you ever need to exchange or replace the antenna, you may wish to do as I have. I made up a section of RG-8 coax. I then dressed and tinned one end and inserted it/them under the feed-point hose clamps. I then coiled a length of the coax into a 5 turn, 5 inch diameter coil (the decoupling loop). I attached it to the (metal) mast as shown in **figure 2**. For the final touch, I fitted the (lower) end with an SO-239 (female) coax connector.

To adjust the Vswr to frequency (145.770 MHz for 2 meters or 51.12 MHz for 6 meters) slide the feed point (hose clamps) up or down as required. You should be able to set the VSWR very close to 1:1 (It is best to make the VSWR setting using at least 25 feet of coax between the SWR meter and the J-pole connector). I have a ten foot section of 1 & 1/4 inch pipe standing in my backyard with three feet of it in the ground. I attach the antenna to the pipe using 2 inch hose clamps. A long length of (not coiled) coax allows me to move away from the antenna at least a wavelength while making the VSWR tests. In short; Do not stand near the antenna while testing VSWR.

Having Fun while saving lives with PacketRadio is what we're all about. For more information and illustrations for the J-pole, visit the PacketRadio Networking home pages on the internet at; http://www.PacketRadio.com/jpole.htm

You will find even more PacketRadio information at:
http://www.PacketRadio.com
or http://www.PacketRadio.org
Visit my personal pages at: http://www.BUXcommco.com
You may send Email to me at: k4abt@PacketRadio.com
73 de BucK4ABT

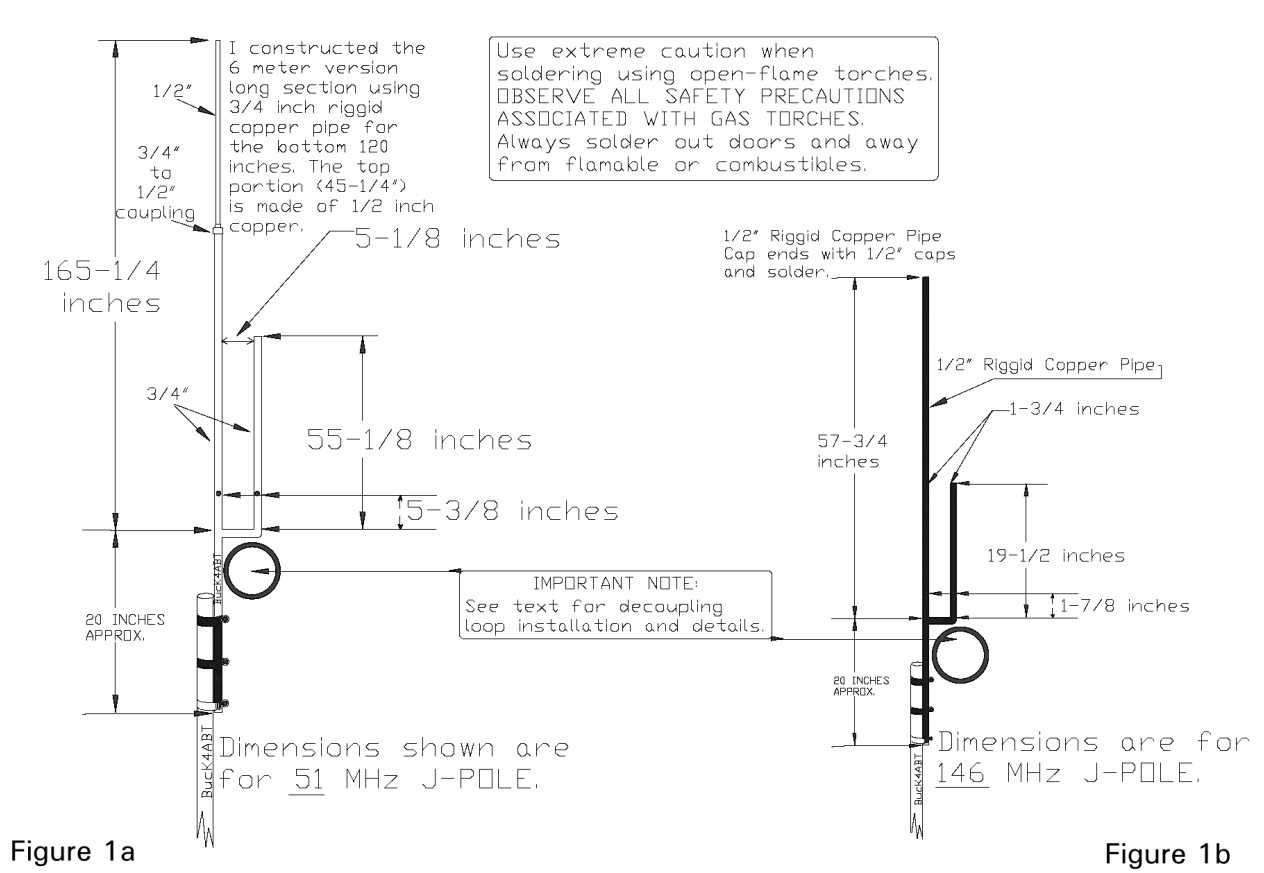


Figure 1a

Figure 1b

Rigid copper water pipe is not as heavy as some might think. Especially when it is only 58 inches long at 2 meters (1a), and when used at 6 meters (1b), a J-pole is about 166 inches long (see text).

To get the correct amount of heat to braze the EMT together, I must use a combination torch/flame that uses oxygen and acetylene gas or a less expensive Benz-O-Matic™ oxygen and MAPP gas.

The latter is available at most Lowe's Building materials for less than 40 dollars ($39.95). The kit comes with enough welding rods to build one antenna. It also comes with two disposable tanks; one of oxygen and one MAPP. The catch is… one canister of oxygen is not enough to finish the job (one 6 meter J-pole). If you ever consider such a project, make sure that you have at least two canisters of oxygen ($7.95 each).
In any case, make sure you are out of doors and away from combustible materials while using this type welding system. ***PLEASE use extreme caution when working with open flame torches of any kind.***

SOME NOTES ABOUT J-POLES:

There are very good reasons why the J-Pole antenna has become so popular in the past few years. They are economical and easy to build. Now that we have the support data, we are finding they are also rugged enough for the more hostile environments that we have at our node and repeater sites, even below 145 MHz... on our 9600 baud backbone at 51 MHz.

The 6 meter J-poles that I've installed on my nodes are exhibiting over 4 db increase in signal strength above the dipoles that were in use on my 9600 baud (6 meter) nodes.

The most recent antenna I replaced with a J-pole was a 2 meter (see figure 1b), aluminum, base station, commercially built, amateur antenna purported to have more than 3 db gain over a dipole. The J-pole I replaced at node K4ABT-7, alias 007 has increased the average signal at fifteen (15) miles by more than 5 db; ENOUGH SAID!

Our J-pole does not have radials! SO WHAT.. well this is what. Without radials, there are antenna (RF) currents that flow along the outer surface of the coax cable shield that introduce VSWR along its length. This kind of RF action tends to corrupt performance of an antenna that has no radials. Two things can happen; RF may be absorbed and hot-spots along the coax shield will raise the VSWR to an unusable level. To prevent this from happening we add a form of decoupling loop just below the feedpoint of the J-pole (see figure 2).

Our decoupling loop is sometimes mistaken for a matching balun. This is OK as it does perform some of the properties of a balun while altering the flow of RF back onto the coax cable shield. In short, this decoupling loop performs by canceling these effects, and creating a balanced to unbalanced matching transformer. Oh, didn't I say that before… the J-pole is a "balanced" fed antenna… the 50 ohm coax that we use to feed it with is "unbalanced," thus we have another reason for adding the decoupling loop… .. OK Lew, … it's a balun.

To make our decoupling loop, for the 6 meter J-pole, let's use a length of the same type coax that we use as the feedline. Make a five (5) turn coil (**4 turns for 2 meters**), five (5) inches in diameter and tape it to J-pole mast, below the feed point(s).

We do this near the point where it connects to the antenna (***see figure 2***). In either case, this five turn, five inch diameter coil serves to isolate the feed line's outer conductor from reflected energy (RF) of the antenna radiating element(s).

To set some of you at ease with the feed point question: I did run some tests on the two ways to feed the antenna…e.g. whether it made a difference if the center conductor were connected to the short tuning stub or longer (radiator) section of the J-pole?

Zig or Zag:

Although there is not a big change in VSWR, let it be known! There is a *pronounced* (***worse***) difference in field intensity when the coax center conducted is attached the short stub of the J-pole; NOT GOOD! Without a lot of fanfare, ... **A RULE**; ***Be sure the coax CENTER conductor is attached to the LONGER SECTION of the J-pole*** and the coax shield attaches to the short section.

SOMETHING TO THINK ABOUT:

All the SNO's who support the Packet habit by supporting a Packet node at a high location knows or has experienced what I'm saying. So while I was atop one my Packet node sites an idea hit me… as Steve N4JTH says… I had a brain.. uh.. "wind-breaker."

Most of the antennas that I have been installing at the new node sites are the copper version of the J-POLE. I've removed most of the fiber-glass antennas and ALL the aluminum composition types and models.

I've learned that aluminum was not intended for use at high elevations, especially where ice and wind wreak havoc on them all winter… or until they succumb to the onslaught of this kind of environment.

I tried the next level of durability in the metallic family; copper. Rigid copper water pipe is not as heavy as some might think. Especially when it is only 58 inches long at 2 meters and when used at 6 meters, a J-pole is about 166 inches long. To make the 6 meter version a bit lighter, and still maintain the durability, I make the bottom 120 inches of 3/4 inch rigid copper, and the final (top section, 45 and 1/2 inches), I make of 1/2 inch rigid copper pipe(see figure 1a). We'll discuss this idea further, a bit later. For the record, I find that EMT makes a great J-pole and I find the same characteristics with 1/2" EMT as I find with 1/2" copper.

LET'S TALK ABOUT THE J-POLE:

The J-pole has, for the most part, been a 2 meter and 70 cm antenna. Matter of fact, the first ones were designed and constructed using 300 ohm, television twin lead (or 450 ohm ladder-line, See Lew McCoy W1ICP J-pole article, JULY 1994). Not many configurations are supportive of building J-poles for frequencies below 144 MHz or for environments that were demanding of a more rugged type antenna.

A few years ago, I replaced a 6 meter **di**pole at one of my Packet node sites (3000 feet) with a six meter **J**-pole (see figure 1a). The first thing I noticed was that instead of just hitting the local 9k6 backbone nodes (those within 50 miles), I am suddenly reaching nodes a hundred miles away.. atop Herndon Mountain in **WEST** Virginia. That is a bit farther than 100 miles. The path is there 95% of the time with the J-pole. The path was there 0% of the time with the old 6 meter dipole.

To help support this data and the use of the more "durable" J-pole, and at frequencies below 145 MHz I received some welcome help at gathering data from **AL (K4ZMC):**

Al Feldman K4ZMC is a long time friend who has been around Packet radio as long as Packet radio has been around. AL installed a 6 meter (copper) J-pole at his "very hostile environment site" above 4000 feet. The node site K4ZMC-7=**MARS** (NO, not on planet MARS.) is near ***Mars*** Hill, NC. Actually, the first antenna went to the 9600 baud backbone node on six meters, K4ZMC-9, alias 9600.

When AL installed the J-pole at node 9600, I had the support I needed to begin compiling data on the performance of the J-pole at six meters… and with Als location being such that it is, it gave support for the durability of the ruggedized version.

Since then, I've begun using three-quarter inch (3/4") electrical metal thin-wall conduit, or EMT as it is called, to construct my 6 meter J-poles. EMT, although a bit heavier, is yet stronger and sustains even stronger winds than the copper version does. I used the EMT version at K4ABT-4=9603 atop Poor mountain SW of Roanoke, VA. The reason I did so here is because I've lost several aluminum antennae at this location, and I felt that I needed to make this antenna a very "beefy" version to handle the high winds and heavy ice loads at this site.

The reason I've not said a lot about the EMT version is because the rigid copper water-pipe version can be built (joints soldered, sweated together) using a single canister, Benz-O-Matic™ propane torch.

This is **not** the case with the EMT version. It must be built using brass compound brazing rods. Thus we have a requirement for a hotter melting flame than a propane torch can produce.

Packet Users Notebook
January 1999

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by
Buck Rogers K4ABT

For association and the record, the ZEPP was one of the early antennas used as a "trailing" antenna behind the Zeppelin (Blimps) of the early forties. Later it was fashioned into an antenna for higher frequencies, and in the seventies, and early eighties the name was changed to the "J-pole" to favor its appearance when constructed as a VHF vertical.

In the late fifties or early sixties, I read an article written by Lew McCoy W1ICP. It was about an antenna called a "Zepp." In his reference the antenna resembled one I had raised called a, "windom." After building Lew's version, I liked its performance better.

Thus began a new era for me, experimenting and building the Zepp (J-pole) antennas. I've since built J-pole antennas that operate on ham bands from 14 MHz all the way to 912 MHz. I've build J-poles of everything from 300 ohm TV twin-lead, 450 ohm ladder line, copper rigid water-pipe and even electrical metallic thin-wall conduit.

As I add new nodes to mountaintops here in Virginia, I often come up with another idea for a future **PACKET USER'S NOTEBOOK** article. For me, finding material for a future PacketRadio column is always easy. There is so much to write about, it's as if there is a never ending supply of fodder just over the next ridge.

This time, I think I have come across something that will be of interest not only to the PacketRadio system node operators, but to the voice repeater operators as well. Especially where the Packet node or voice repeater is situated on a mountain top that is a weather hostile location during the winter months.

Adding a new PacketRadio node or voice repeater is only half the battle. When I talk about "hostile environments," I am referring to the torture that our node or repeater antenna is subjected to during the heavy ice loads, snows, winds, and rain. When we are warm and comfortable in our home, the node or repeater site high on that mountain top may be receiving gale-force winds and the antenna may be slowly loading up with a new layer of radial ice.
My friend Chuck and I were atop one of these mountain tops recently and the wind was coming at us like a cold … no, a frozen knife. The gusts were topping over 50 mph. Now to some, this is not too heavy a wind, but when the wind continues at this clip, hour after hour, the toll on the site usually begins to show signs of wear and tear early on.

The part of our site that takes the worse beating is the antenna. I've tried almost every kind, shape, and model at these sites. At 3000 and 4000 feet, we begin to see damage to the antenna soon enough.

I began thinking; If this is bad here on this lil 3000 foot knob, then ole Fred WB4QOC really has a problem with his node antennas. atop Grandfather mountain, NC (nodes WB4QOC-5 and WB4QOC-6) and Mount Mitchell, NC (nodes WB4QOC-2 and WB4QOC-3). Grandfather mountain is over 5400 feet and Mount Mitchell is the highest point in the USA east of the Mississippi river at 6684 feet.

We do know that winds in excess of 190 mph have been recorded atop Grandfather mountain, but we understand that the equipment used to measure the wind velocity at Mount Mitchell was blown away… pegged out, and burned out the bearings in the anemometer. The first winter that node 6684 was active, he replaced the ISO.. something-another with a stainless-steel whip. It lasted the rest of the season… until it was found in the parking lot of a nearby lodge. It was shot!. Our next try was with a very stubby, quarter inch diameter Wintenna dipole. Dern that thing lasted through the next winter. It must have been a good one… someone climbed up there and took it. We know it was not blown away… wind don't use cable cutters to remove antennas. I think Fred went with a insulated chunk of water-pipe. That is until now.

Design Your Own J-Pole Antenna

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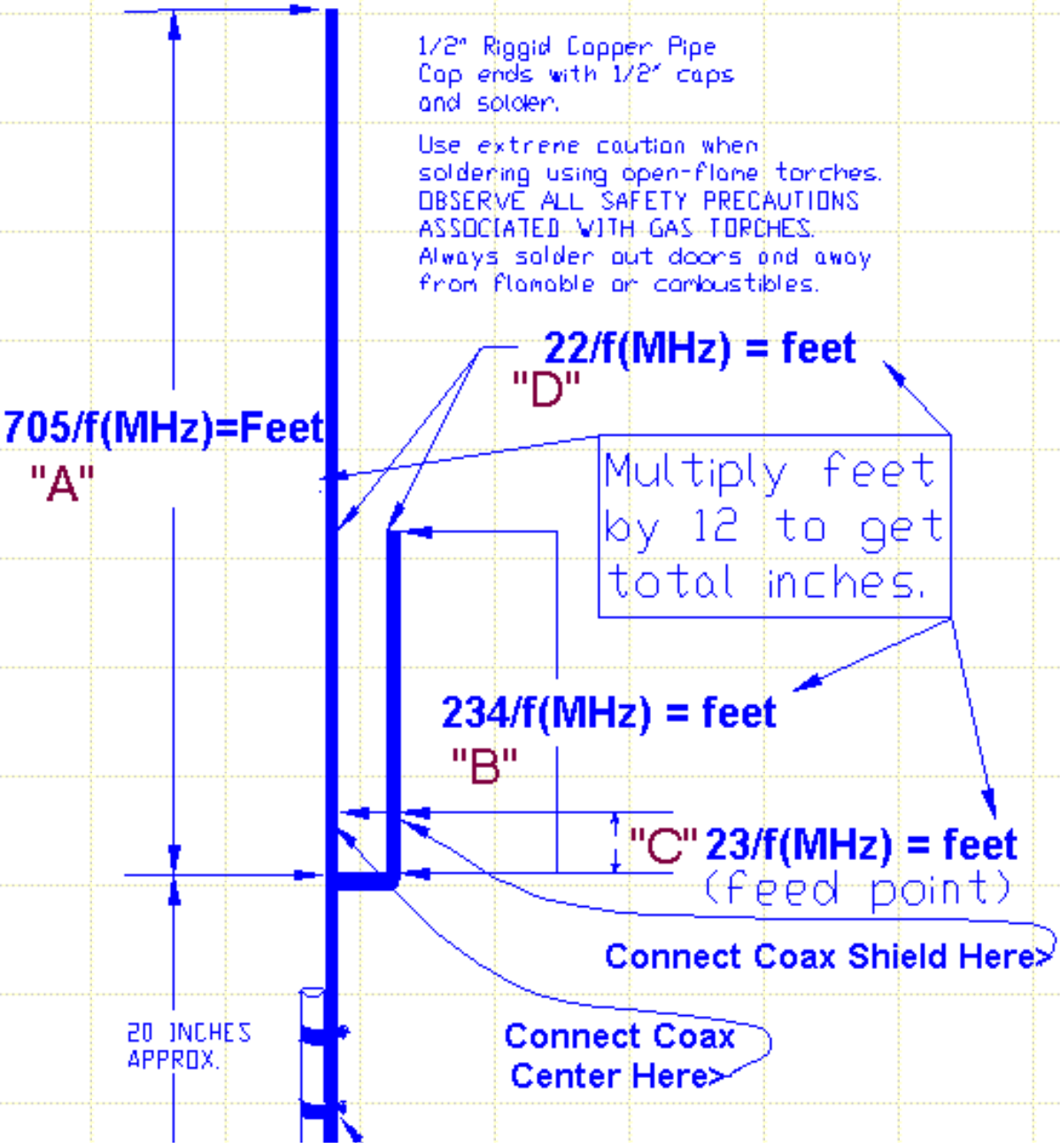
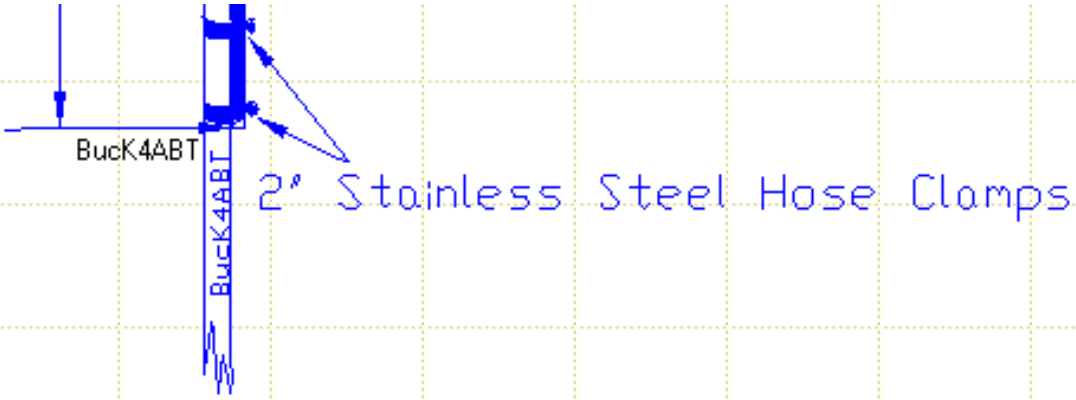




SAVE TIME BY USING THE HANDY J-Pole CALCULATOR BELOW.

NOTE: Coax center conductor attaches to the "Long section" feed point.
Shield attaches to the short section feed point.

Enter Antenna Operating Frequency

Freq MHz

A (Long section) dimension is:	***feet,***	***inches,***	***meters***
B (Short section) dimension is:	***feet,***	***inches,***	***meters***
C (Feed point) dimension is:	***feet,***	***inches,***	***meters***
D (Spacing) dimension is:	***feet,***	***inches,***	***meters***

*Inside (spacing) dimensions are metal to metal measurements, **NOT** center to center.*

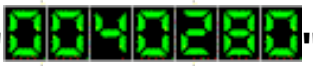
RASCAL PSK Interface

Click here to view an alternate method for feeding a VHF copper "J-Pole"

CLICK HERE for a "close-up" of the alternate J-Pole feed method.

©1996-2004 G. E. "Buck" Rogers Sr.

WebMaster, Buck Rogers K4ABT

Over Five Million Visitors "  " since January 2001

Back to the ***SEDAN*** Antenna Designers Page

Go To the ***BUX CommCo*** Home Page

Go To the ***PacketRadio*** Home Page

" This is Walter... Communication is dubious... I may have heard you, or your phantom transmission... I will attempt to reach your shadow and raise you..."

(Thanx Firesign Theatre)

THE LOW BANDS

If you like DXing on the low bands, 160, 80, 40, and 30 meters, these sites may be of interest- they include info about both TX and low noise receiving antennas and techniques. Enjoy!

KB1GW Beverage Antenna Collection	Funkenhauser MW DX Links	W3LPL on Beverages	Who's Who on Top Band
	Reversible 40m Delta Array	Flag & Pennant Antennas	K3KY's Short Beverage
	NEW FT-1000D Diversity Rx	The K6STI Loop	Bobtails and Half Squares
High Band Bobtails/Half Squares	NEW The G3TKF Synchro Mod	The K9AY Loop	W7AV Tries a 40m Half Square
Top Band Reflector	NEW More Links- FT-1000D Mods	Bobtails W4RNL	Half Squares W4RNL
NEW G4VGO 160m Kite/Balloon Ants	NEW EI7BA 160/80 Antenna Ideas		

EMAIL Send Comments to K3KY

Ham 'Treasure' Sites PREV

HOME K3KY's DX Toolbar

About K3KY NEXT

K5DKZ

Antennas

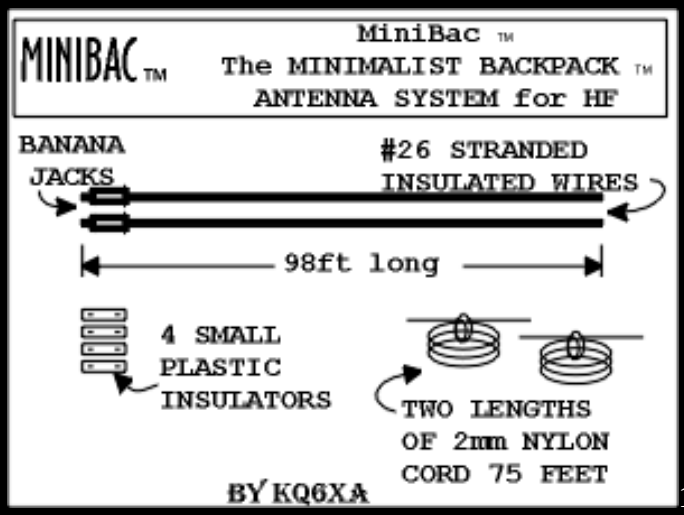
[Short Dipole](#) - [Efficient Vertical](#) - [Beams](#)
[VHF Discone](#) - [Trap Dipole](#) - [40m Beam](#)
[Shunt Feeding a Tower](#)
[Some Basic Antenna Stuff and Ideas](#)
[AM Broadcast Band Loop](#)
[Multiband Antennas](#)

THE MINIMALIST BACKPACK ™ ANTENNA SYSTEM

(C)1999, 2000 BONNIE CRYSTAL KQ6XA

HF Antenna for the Micro-Light Backpacking Enthusiast

SIMPLE CONSTRUCTION DETAILS AND SUGGESTIONS ON VARIOUS FIELD CONFIGURATIONS



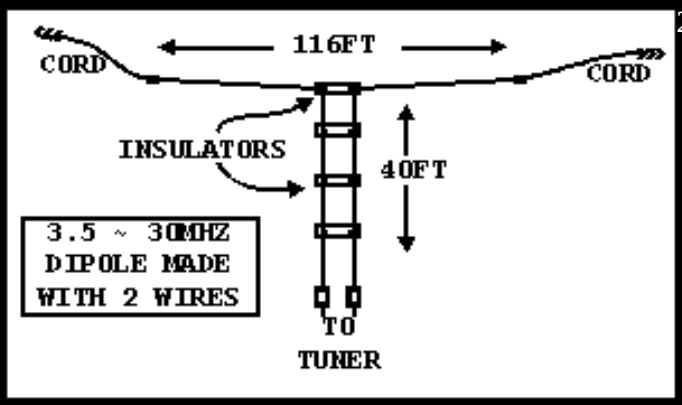
1

MiniBac.

The Minimalist Backpack Antenna System. Designed to be extremely efficient for both weight and radiation.

Components of the Minibac

QTY	ITEM
2	98 FEET, STRANDED COPPER INSULATED WIRE, #26 AWG
2	75 FEET, NYLON CORD, 2mm
4	INSULATORS, LIGHTWEIGHT
2	BANANA PLUGS

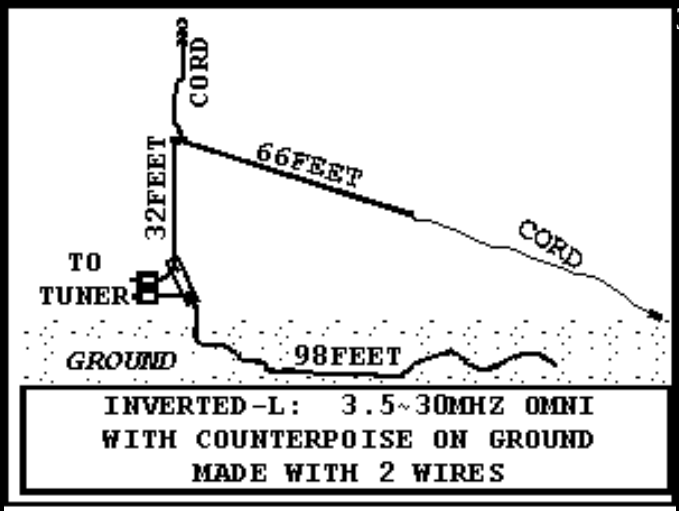


2

Virtually unlimited configurations of the MiniBac are possible. Miniback forms its own feedline, which is quickly assembled in the field to meet the requirements of your site and choice of style. No lossy, heavy coaxial cable, or bulky twin lead is used. Open wire feed makes every precious watt count.

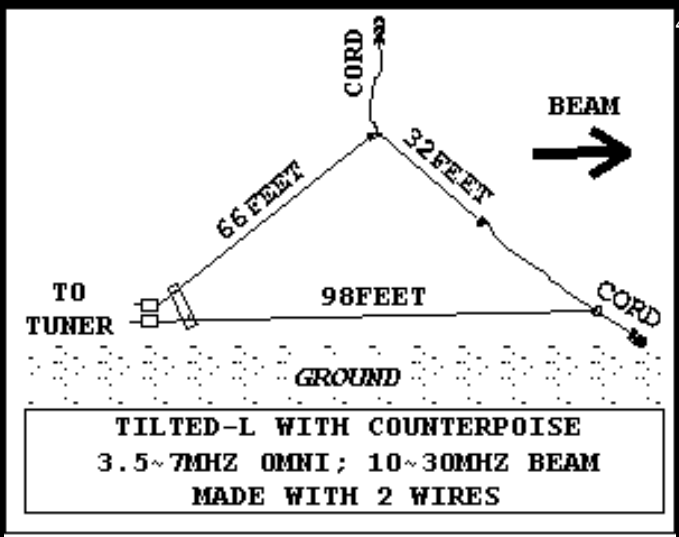
<#2-LEFT: The MiniBac forms a multi-band dipole using 2 supports, or an inverted-Vee with a single support.

3



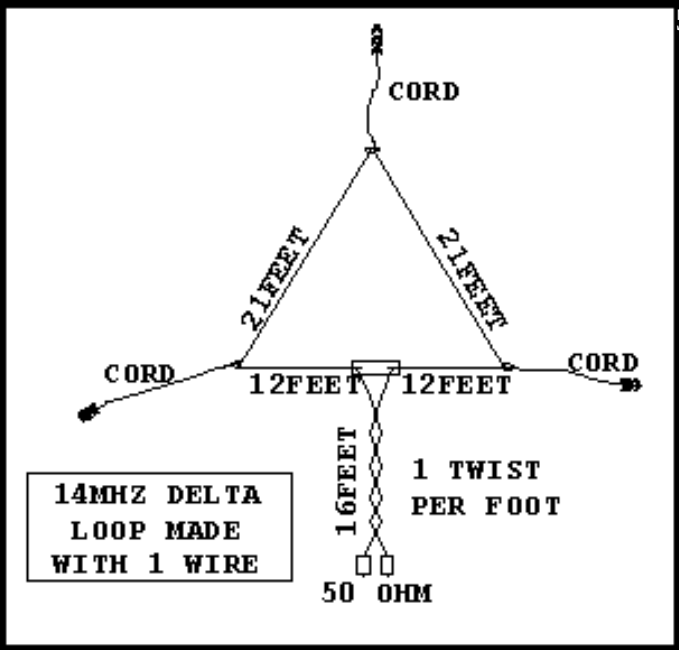
<#3-LEFT: Using a single support, such as a tree or pole, the MiniBac forms an Inverted-L with efficient counterpoise simply laid on the ground using no special alignment. This antenna configuration provides a good omnidirectional pattern, with both low angle and high angle radiation. While a tuner is suggested for multiband operation, no tuner is needed for 50ohm operation on 7MHz.

4



<#4-LEFT: Using a single support, such as a tree or pole, the MiniBac forms a Tilted-L with counterpoise raised above the ground several feet high. The raised counterpoise method provides slightly better radiation effiency. This antenna configuration provides optimum low angle signal on 10 through 30MHz as indicated in the direction of the beam arrow. On 3.5 through 7MHz, it is omni-directional with high angle radiation for communications within about 350mile radius.

5



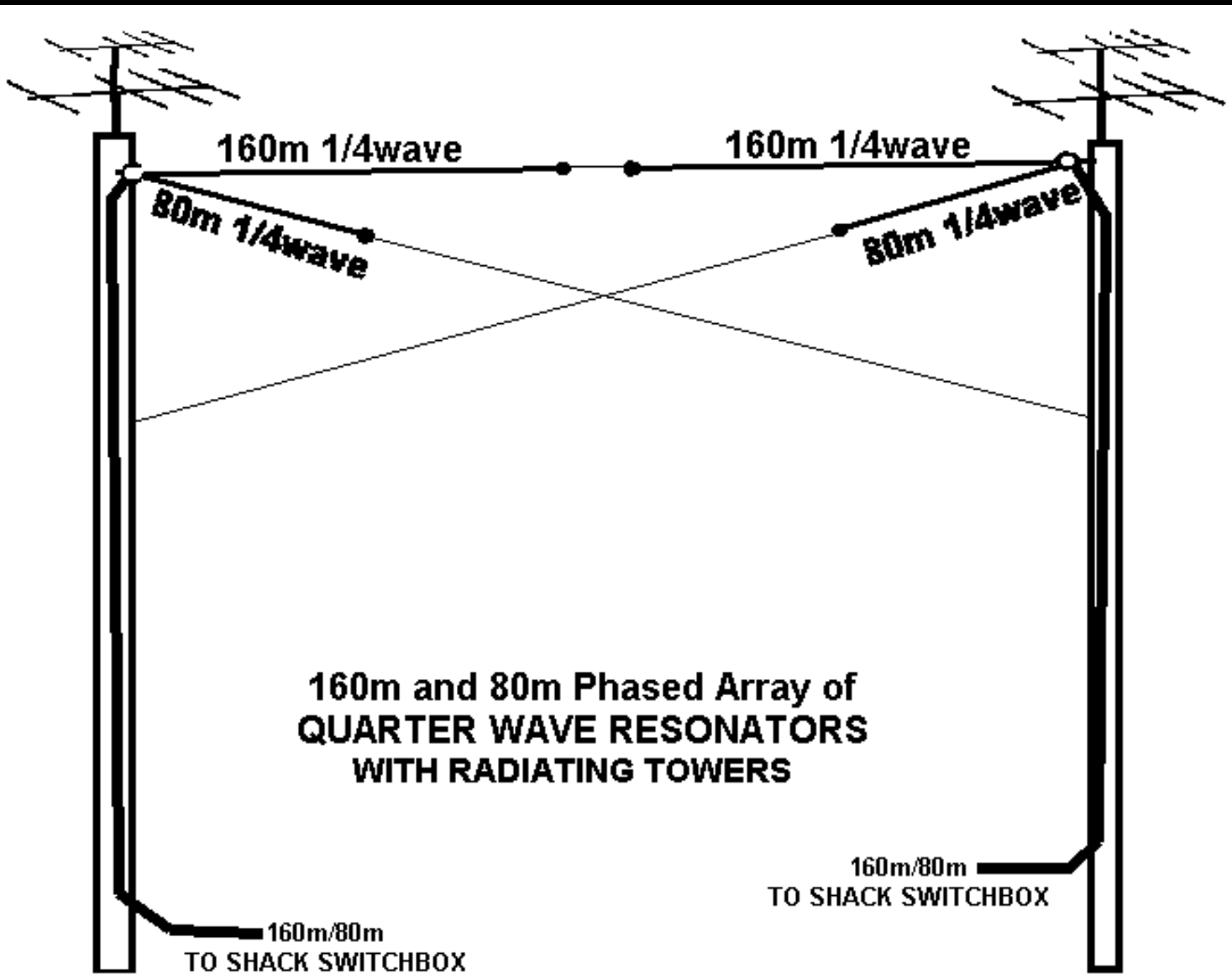
<#5-LEFT: A single wire of the MiniBac forms a Delta Loop antenna strung from a single support tree or pole. This method provides high efficiency with good directivity on 14MHz, requiring no tuner for 50ohm match. Using the twisted pair feedline, which can be run just above ground level, this configuration may be used with good success on other frequency bands, 18 to 30MHz, but a tuner is recommended. In addition, for 3.5 and 7MHz operation, the remaining MiniBac wire can be laid on the ground as a counterpoise, similar to the #3 Inverted-L, with the Delta-Loop and its feedline used as a vertical radiator against the counterpoise.

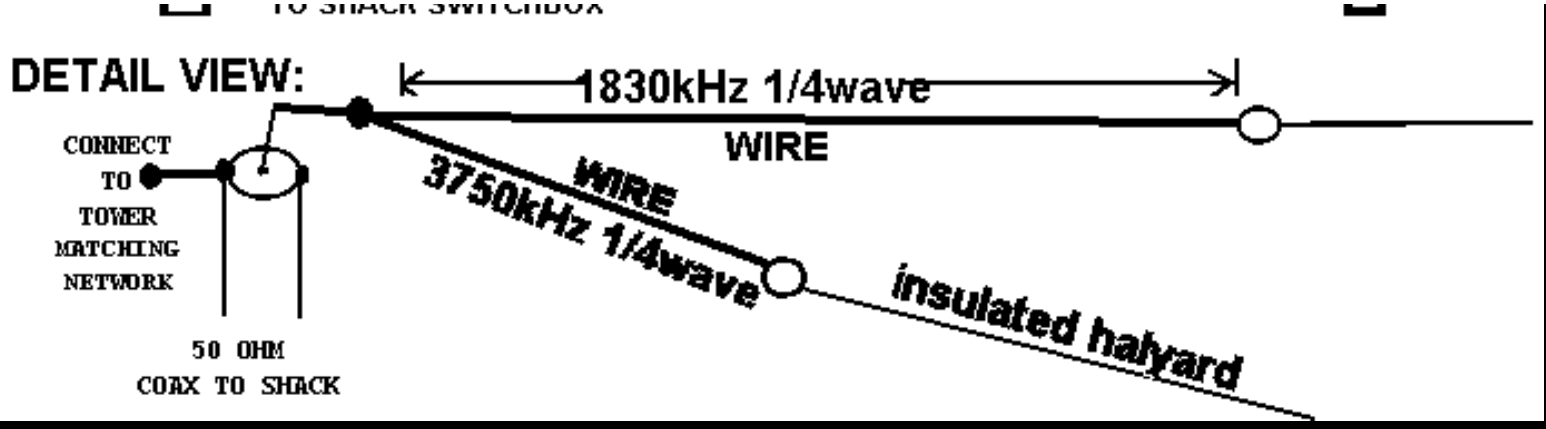
Ham radio operators are welcome to homebrew and use the MiniBac for personal and club use. Non-commercial use of the MiniBac name for ham radio in connection with the antenna is authorized. All commercial rights are reserved. Links to this site should direct to the home page. Permission for use of drawings should be requested by email.

Other Interesting Antennas by Bonnie Crystal KQ6XA:

Broadband/Multiband Dipole and Tower Radiating Antennas for 80m and 160m
Using two towers spaced about 250feet apart --BONNIE KQ6XA

#1: Phased Array of Radiating Towers using Multiband 1/4 Wave Resonators for 160m/80m





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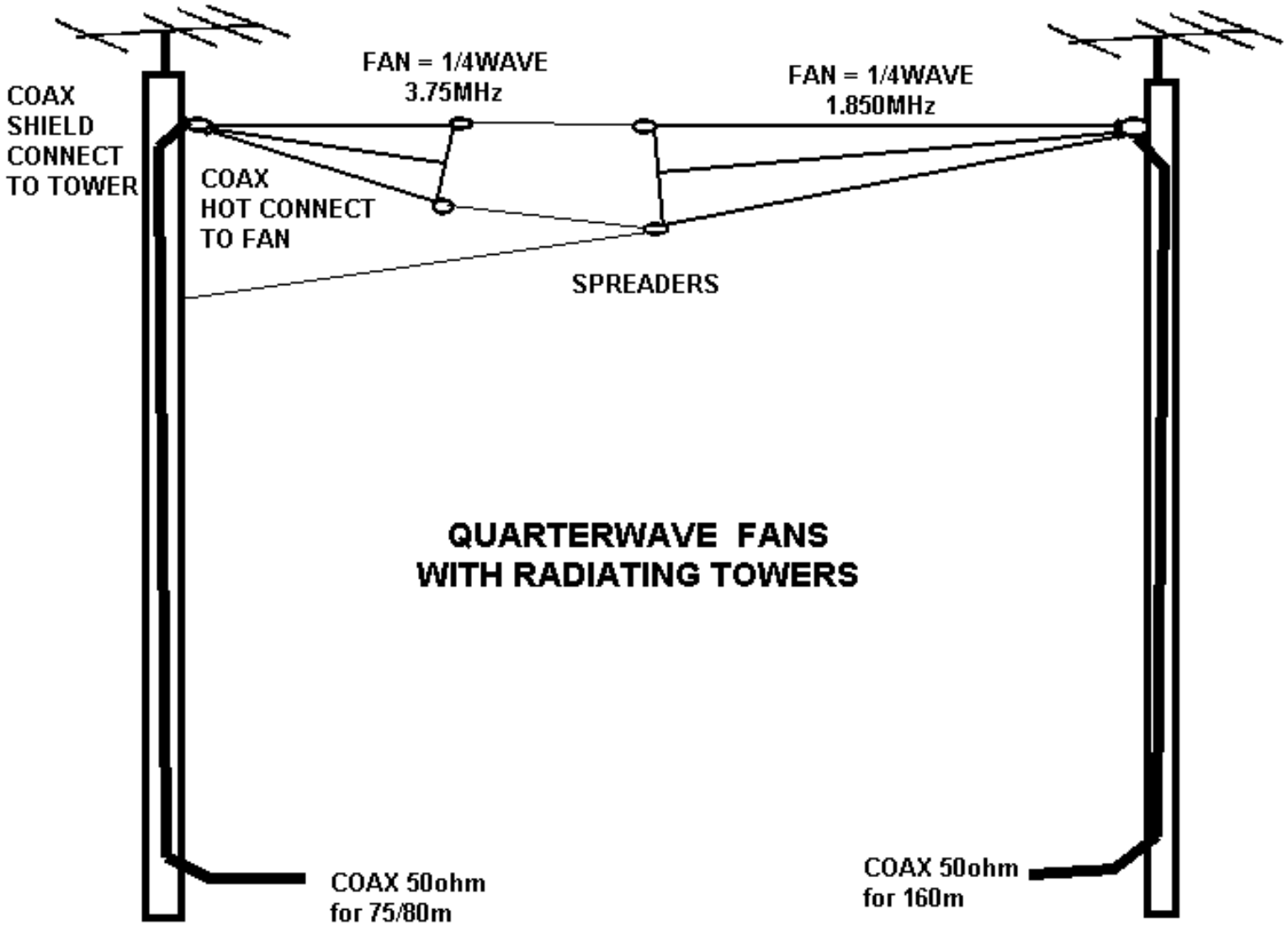
#2 Radiating Towers using Broadband 1/4wave Fan Resonators for separate 160m and 80m



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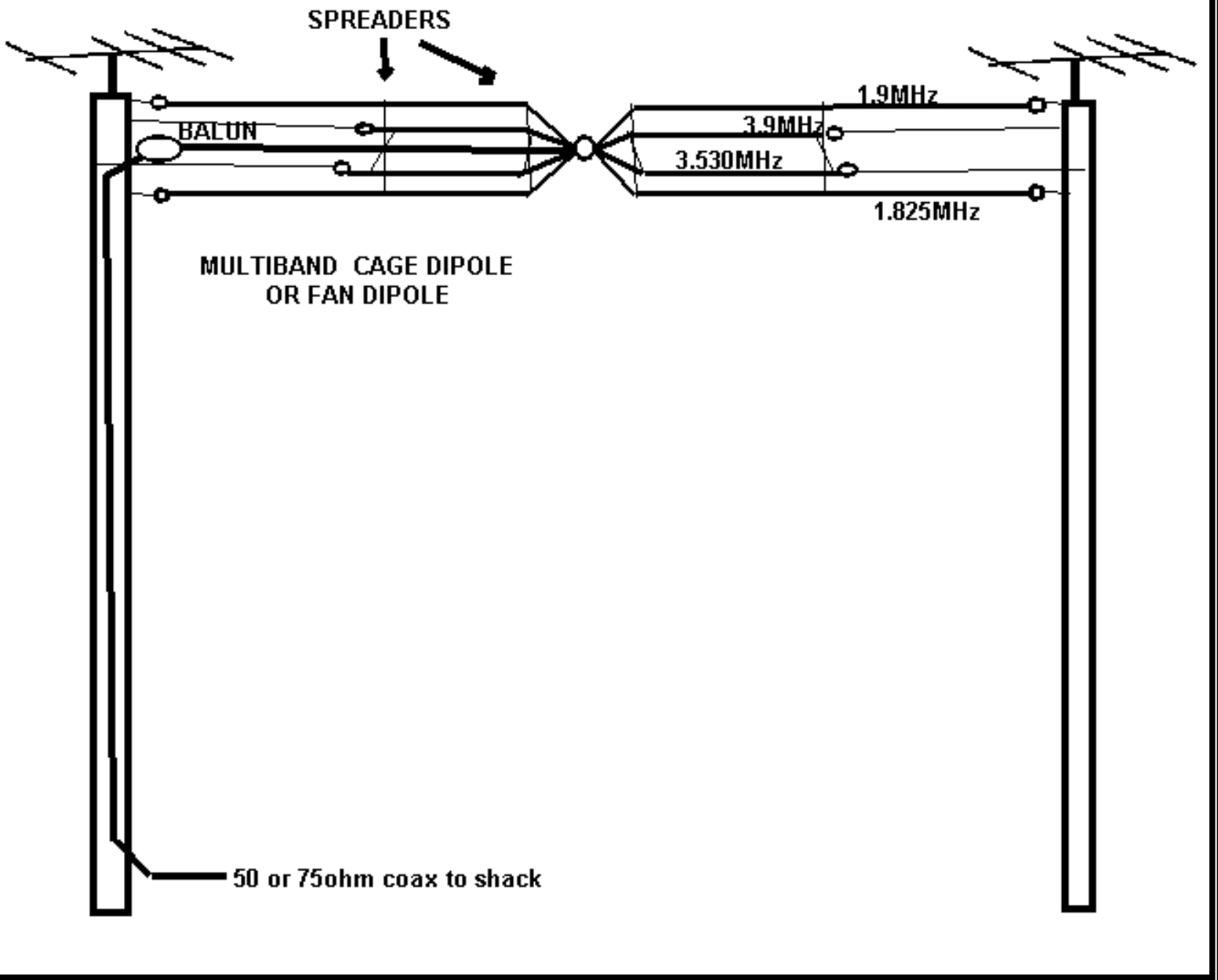
#3: Multiband Cage Dipole for 160m/80m



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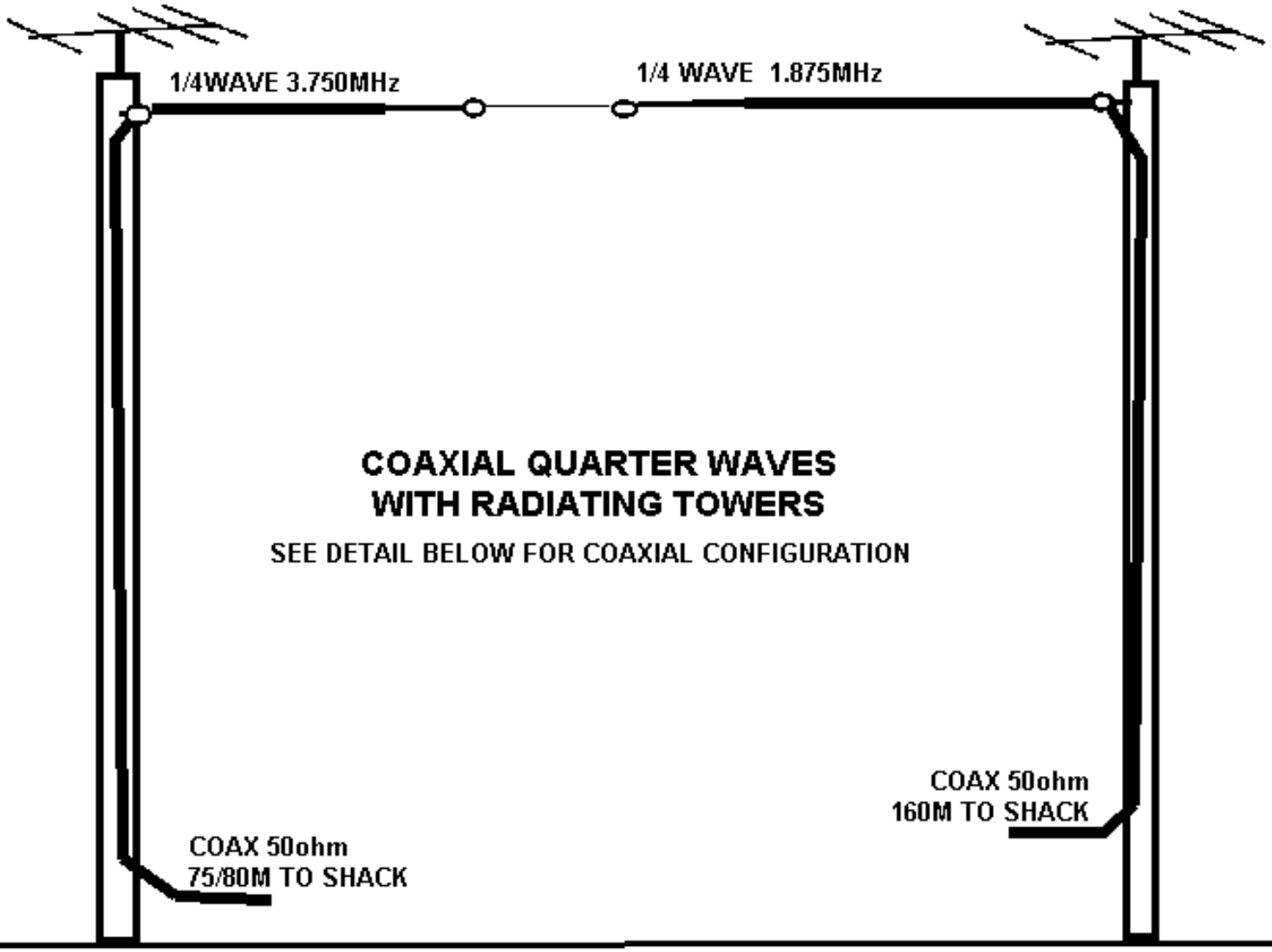
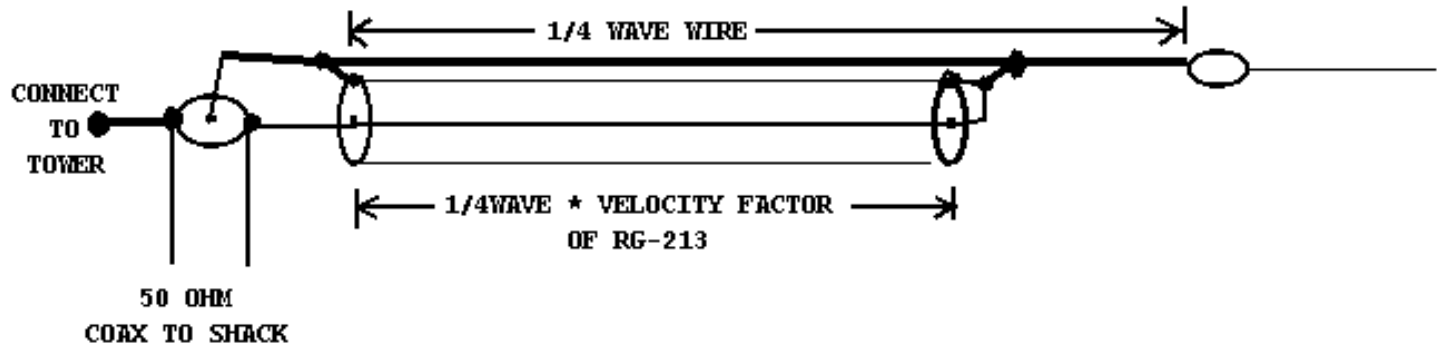
#4: Coaxial Bazooka 1/4 Wave Resonators with Radiating Tower for separate 160m and 80m





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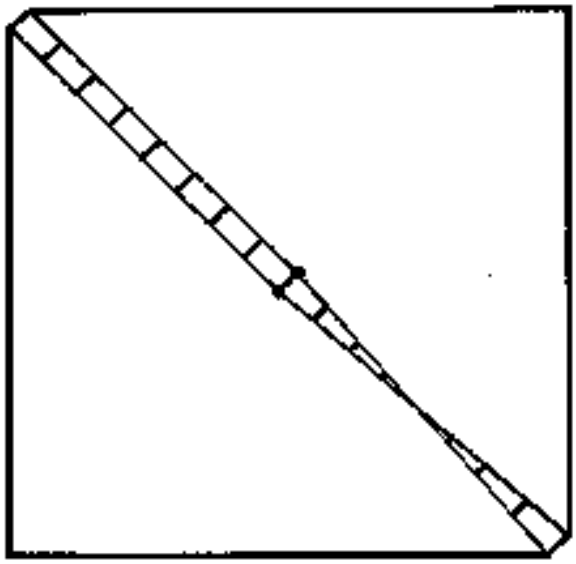
[CLICK HERE TO GO BACK to the KQ6XA HOME PAGE]

The K6STI Receiving Loop

This relatively unknown antenna does an amazing job pulling weak low band signals out of man-made noise backgrounds

DX SECRET WEAPON



Although it was prominently featured in September 1995 QST Magazine (p.33-41), I believe this antenna has been overlooked by a lot of low band DXers. Perhaps it seems a little too complicated, but I found it well worth the effort it took to design and build one. It is a terrific performer and has greatly exceeded my expectations. This is a low, medium size, horizontal loop with some *very* interesting characteristics. It has an omnidirectional, fairly low angle pattern with a deep overhead null like a vertical, yet it rejects vertically polarized RF energy and responds only to horizontally polarized energy. It offers a 25 to 30dB improvement in signal-to-noise ratio over a vertical for man-made noise. This is often the predominant noise type in densely populated urban and suburban areas. Results often seem magical as signals that are covered by noise on a vertical antenna become perfectly readable upon switching to the loop. A typical antenna is about 21 feet (6.5m) on a side and about 10 feet (3m) high. (Mine has 23ft (7.0m) sides) Height is not important as with Beverages- the antenna works just as well at 5 feet or 50 feet (1.5-15m). It is double-fed with ladder line in diagonally opposite corners to maintain good balance. The feedpoint is at the center of the ladder line. Coax from the receiver is connected through a matching network with a transformer and series resonating capacitor in the secondary, built in a weatherproof enclosure. For a 21 foot loop, there is a small impedance stepdown on 80m and a larger ratio on 160m. The antenna can be built with remote band switching, as I did with mine. It works well on 80 and has been surprisingly good on 160 (I had not expected much.) It is often my best performing low noise receive antenna.

Low band operators who have already used Beverage antennas are nearly always disappointed with the K6STI loop. Beverages will 'spoil' you for any other low noise receive antennas, but many hams do not have room for them. If you want to reduce atmospheric noise and you do have some room for Beverages or phased verticals, both of which are directive, then those antennas would be a much better choice. Likewise, the WA2WVL cardoid loop array provides good directivity over a narrow bandwidth if you have the room- about 80 feet on 80 meters. (QST, August 1993, p.31) The Ewe, K9AY, and Flag and Pennant antennas also provide good directivity in a fairly small space, but with lower receive signal levels. They might benefit from the use of a receive preamp, and so might the STI loop, to a lesser degree. (I do not use a preamp with mine, however). It is a good idea to add a

highly selective tuning network ahead of any preamp to avoid receiver front-end overload. This loop has an omnidirectional pattern and, in theory, it should provide no improvement in S/N for atmospheric noise. K6STI and W6KUT, who built the first loop, noticed an apparent improvement of 1 to 2 S-units even when there was no detectable power line noise present. I have also seen this behavior in my antenna. STI reasons that in populated areas the noise background may consist of many individual man-made noise sources which, in composite, are characterless. Anyone interested in this antenna is encouraged to get hold of the QST article and carefully read and reread it for full understanding. Topband operators might want to consider building a double size loop around 50 feet (15m) on a side if it will be used only on 160m. Matching network component values would have to be adjusted accordingly. Because I do not have room for low band Beverages, the K6STI loop has proven to be a most valuable low noise receive antenna for me.
(Revised and updated 8 Apr 2000)
(Second update 13 Feb 2002)

W7LR tried a 50ft version of this antenna on 160m. He used a "pre-war" Terman handbook (p. 52) to determine that the loop inductance on 160m is about 110uH. He calculated the resonating capacitance at 1830KHz to be 60pF. His transformer was 4 turns, antenna side, and 20 turns for 50 ohm impedance coax. The core was an FB77-1024 ferrite bead. He said it "tuned nicely" (actual resonating capacitance was close to the predicted value). There were no comments as to the actual antenna performance, only that he has tried a lot more receiving antennas since this September 1995 design.

EMAIL Send Comments to K3KY

The Low Bands PREV

HOME K3KY's DX Toolbar

NEXT

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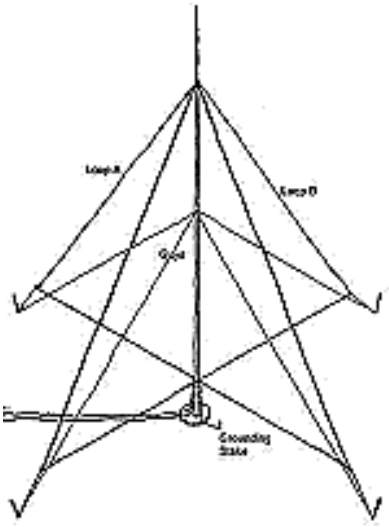
hcdx k9ay loop

K9AY is a loop developed in the early 1990's by American radio amateur Gary Breed (K9AY).
Since then DXers, ham operators and comerical enterprises, all have raved and ranted about this loop, calling it "miraculous", "stunning" and the antenna that beats anything.
While comercial K9AY can be bought at a price, this loop can easily be homemade at low costs.
Here we tell you all about it.



Basic info

K9AY compact loop for lower bands

Gary Breed, K9AY, has taken loop theory farther and describes an antenna that can be erected in almost any back yard and provides performance better than most antennas. *-- Read more --*

Gary Breed's K9AY description [PDF]

Comments on the K9AY terminated loop [PDF]

K9AY, *the* antenna for serious MW DXing

Bjarne Mjelde: This is without doubt *the* antenna for those who want to do serious MW DXing and do not have the space to erect multiple long beverage antennas. Its gain makes it very effective for any kind of DX – be it nighttime, greyline or daytime. *-- Read more --*

Wellbrook's comercial K9AY system

Dave Kenny, British DX Club: After some four months of using the Wellbrook's comercail K9AY I find that I am completely hooked and don't know what I'd do without it.
-- Read more --

Homebrewing

First K9AY for medium wave

Al Merriman, K4GLU, was the first to use this antenna for serious medium-wave DXing. A description of this loop. *-- Read more --*

More on K9AY

Basics
General info on the K9AY loop

Performance
Listening test and other observations

Homemade
How to make your own K9AY

Grounding
The importance of good ground

Wires
How to place wires best

Remote
Remote control of the KPAY

- Front page
- DX News
- Logs
- Andes DX
- Antennas
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▶▶ Build your own K9AY

The performance of the K9AY is miraculous, says *George Maroti*. Then gives details on how to construct one yourself. -- *Read more* --

▶▶ Connelly's loop version

Mark Connelly's version, with drawing, of a K9AY loop. Also describes a scaled-down version of the loop.

-- *Read more* --

▶▶ One step further: two loop system

Having adapted the original ham K9AY loop, *Al Merriman* took the antenna one step further, by setting up a two K9AY loop system.
-- *Read more* --

▶▶ Two loops working as four

Johan Bodin: I think I have invented a way to make two loops appear as four loops... -- *Read more* --

▶▶ Going even further: 8 loop system

A system with 8 loops definitely works, but there are situations where nulls are difficult to get. -- *Read more* --

▶▶ Modifications to the K9AY loop

Six suggestions to make your K9AY loop work even better.
-- *Read more* --

▶▶ More on loop modifications

Single loop will work fine. Also proof for the need of a 9:1 matching transformer. What wire length is best for medium wave. -- *Read more* --

▶▶ Make yourself a protable K9AY

It takes some time to build, but after that you can have your setup up and running in 30 minuter. -- *Read more* --

▶▶ Make your own transformers

Make your own broadband transformer for the K9AY; or buy one ready made from Mini-Circuits. -- *Read more* --

▶▶ Make your own Vactrols

Two ways of producing your own Vactrols for use with your homemade K9AY. -- *Read more* --

▶▶ Check for worn out Vactrols

George Maroti: How to check for worn out Vactrol, without removing the Vactrol from the circuit board.. -- *Read more* --

---

Performance

Stunning K9AY loop test

On a couple of dozen MW signals from Japan and the Koreas, the beverage *never* out-paced the loop. -- *Read more* --

Saviour in the noise

Werner Funkenhauser: I had almost given up on DXing, due to a high noise location. Then came my saviour, the K9AY loop. -- *Read more* --

K9AY compared with Beverages

Martin Elbe: The K9AY turns out to be an excellent performer, able to beat the Beverages in some cases.
-- *Read more* --

Why doesn't K9AY work for me?

Discussion on how to make the K9AY loop work as intended.
-- *Read more* --

Grounding

Important with good grounding

The K9AY antenna doesn't work without ground. It needs *really good* ground to performe optimal. -- *Read more* --

Use many ground rods

Add many more ground rods, and you will get increased gain.
-- *Read more* --

How to balance your antenna

Tom Rauch, W8JI: Discussing the need, or not, for shielding metal boxes, and how to balance different antennas as Beverage, small loops, Flag and Pennant.
-- *Read more* --

Wire matters

Loop wire separation not needed

Andy Ikin, Wellbrook: There is no real need to separate the two loops. In fact the two loops can be shorted together at the apex with no measurable affect. -- *Read more* --

Shape of the K9AY loop

Guy Atkins: While the K9AY is a crossed-X delta loops, other wire shapes may give good results. -- *Read more* --

150 feet should cover medium wave

Antennas up to 150 feet in length, and maybe just a bit more, should

give full medium wave coverage. *-- Read more --*

▶▶ Twin wires gives more gain

Andy Ikin: Experimenting with two wires shows a greater gain. More experiements needed to fully explain why. *-- Read more --*

▶▶ Height above ground to be determined

Andy Ikin: I think some experimentation is required determining the distance above ground of the lower part of the K9AY loop.

-- Read more --

Remote controlling

▶▶ Variable termination an absolut must

Variable termination using a Vactrol or some other method is an absolute must. Peak performance cannot be realized with a fixed resistor. *-- Read more --*

▶antennX ▶Cebik ▶FM antennas ▶Werner's links ▶Antenna Elmer ▶Coax basics ▶DX-Tools.com

Kites and Aerials

[Home] [Up]

Click ***HERE*** for the Amazing **Online Ham Radio Flea Market**

Kites and Aerials

Choosing your kite

6.7m fibreglass poles for sale click ***HERE***

Search G3CWI's site

I'm the world's worst kite flier. It's official. I spent most of my childhood running up and down beaches with sticks, cloth and string. I had not got a clue and neither, it seems, had any of my aunts, uncles and other relations. Perhaps it runs in the family?

So, why am I telling you this? The reason is simple, I reckon that I'm uniquely well qualified to talk about the subject. After all, if I can get a kite to work then you will be sure to be able to get it to work better!

My involvement with kites for radio started in 1981. I was getting ready to go off to Antarctica as VP8ANT. Being a great planner, I was thinking about how to make sure that I could get on the air from there and kites came to mind. I think it may have been due to reading that Antarctica was the windiest continent. Anyway, either that year or the year before there had been an article in QST about a new kind of kite. It was called a parafoil. A company called Jalbert Aerotechnology in Boca Raton (?) Florida was making them. I sent off for some information and they sent me a brochure. These kites were not cheap but they did look promising. They have several special characteristics that make them good for aerial work:

- they are very stable in flight
- they fly at a high angle
- they are light and easy to fly
- they have no rigid parts whatsoever to get broken.

The traditionalists use box kites (I think Marconi started that off) but I knew that I would never get one of them into the air!

My kite arrived and I set about testing it. The amazing thing was that I could launch and fly is easily - I have NEVER had to run with this baby. And what's more it performed exactly as the sales literature said. Bliss.

After a while I got to flying it in stronger winds and then on one sad day, the 100 pound breaking strain line broke. Off sailed my kite, landing in a field of wheat. There was no chance of finding it. I wrote to Mr Jalbert and explained that I was hoping to take the kite to VP8 and that his recommended breaking-strain line had broken. By return, he sent me a new kite - no charge. That's good service!

Anyway, some months later, the wheat field got harvested and the harvester driver actually spotted the kite and retrieved it in tact. That was 1981. I have both kites here to this day.

Don't choose a kite that is too big. Mine seems to be about the right size for most aerial applications. The bigger ones will be impossible to handle in anything but a light breeze. Bigger is NOT better here!

So here is the first lesson. Spend some money on a really good kite (a Parafoil) and it will last for a long time and do great service.

Antarctica

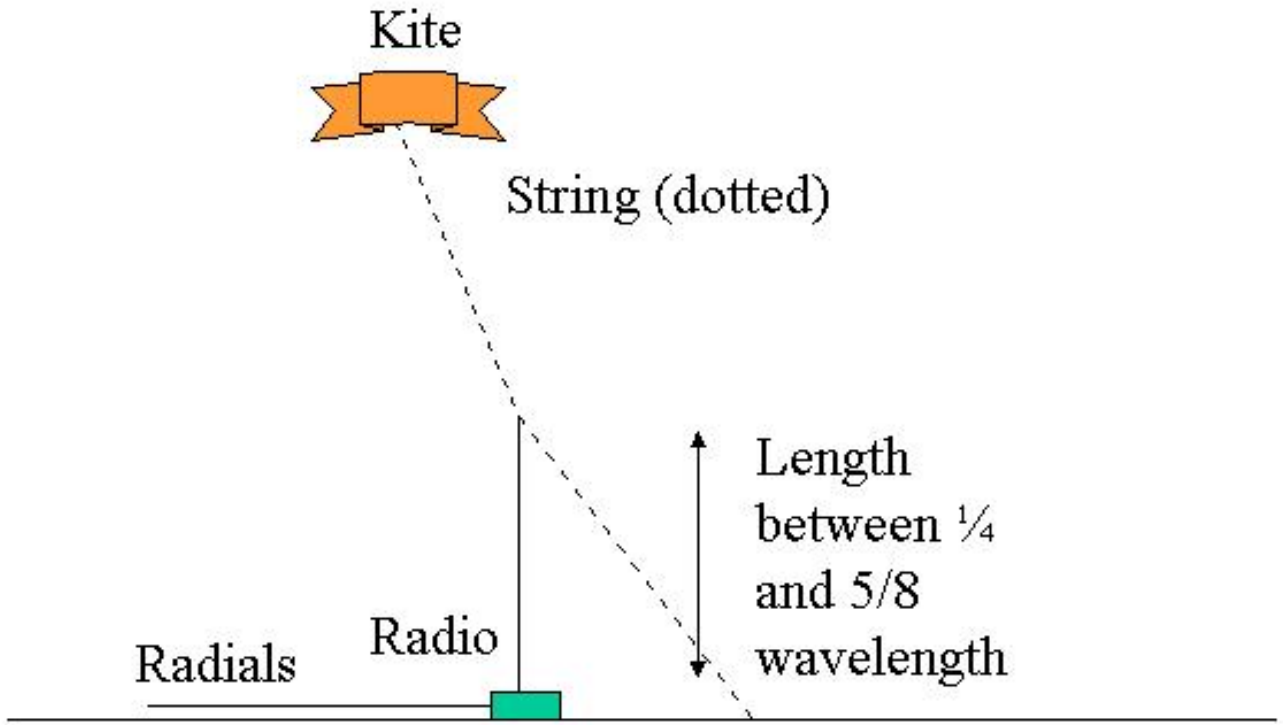
Suitcase packed, off I headed, with my kites in tow. Boy was it windy. I set about trying to break the endurance kit flying record. It was a windy day and I managed about 18 hours before the wind really got up and the string broke again. However, there was not a wheat field for some distance so I managed to get the kite back. Parafoils need the tension of the string to fly. Once the string breaks the fly like a stone. Quite a handy characteristic.

On a field trip. I launched the kite from my sledge and called the base. they were rather surprised to get a call during the day but the kite certainly enabled me to put up a good aerial. I have a picture of this event which I must scan in some time.

I also used the kite to launch an aerial to make some QSOs on 20 metres - reaching back to the UK with just 8 Watts of SSB. So the kite was a success.

Aerial types

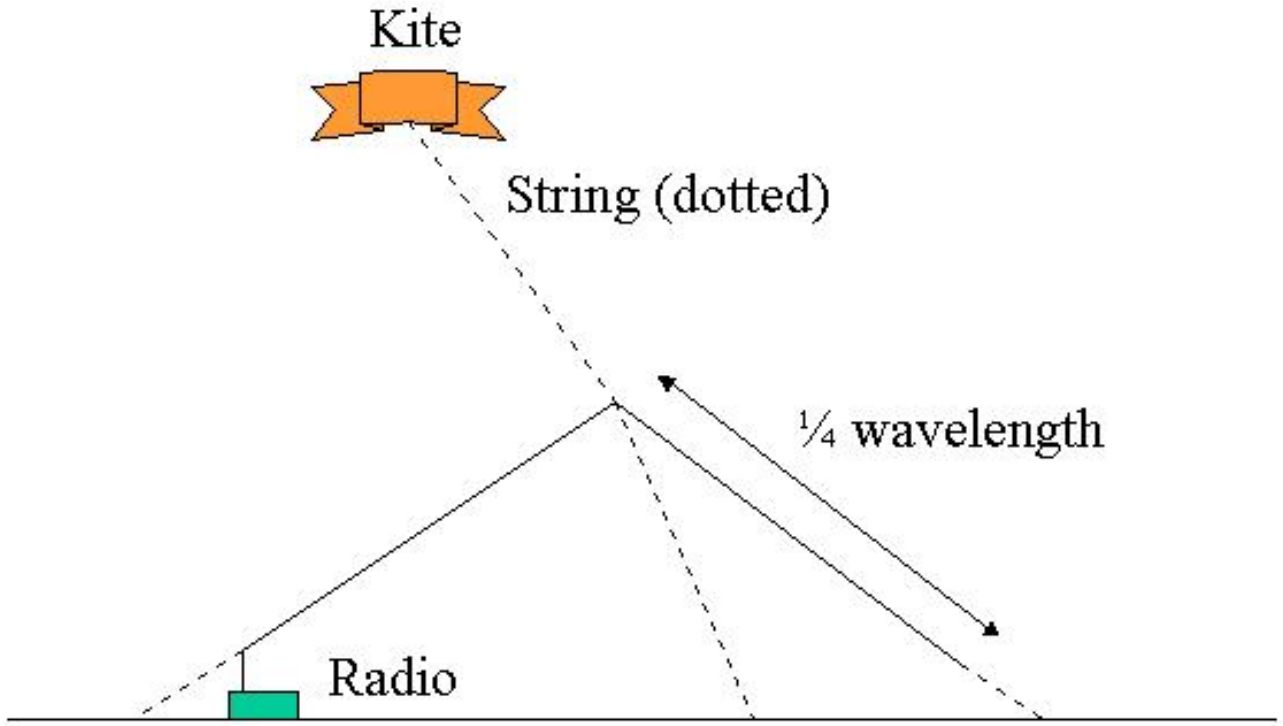
There are several choices. The most obvious is to use an end fed wire as the "string" for your kite. However, as the wind varies, the angle and direction will change (often a lot) which can be undesirable. I tend to hang a wire of the string, about 100 feet down from the kite. This gives a vertical aerial that is much more stable. Add a few radials and a tuner and you're in business.



With two kites, you can be more ambitious and I have flown a large inverted L before. However, my 80m 5/8

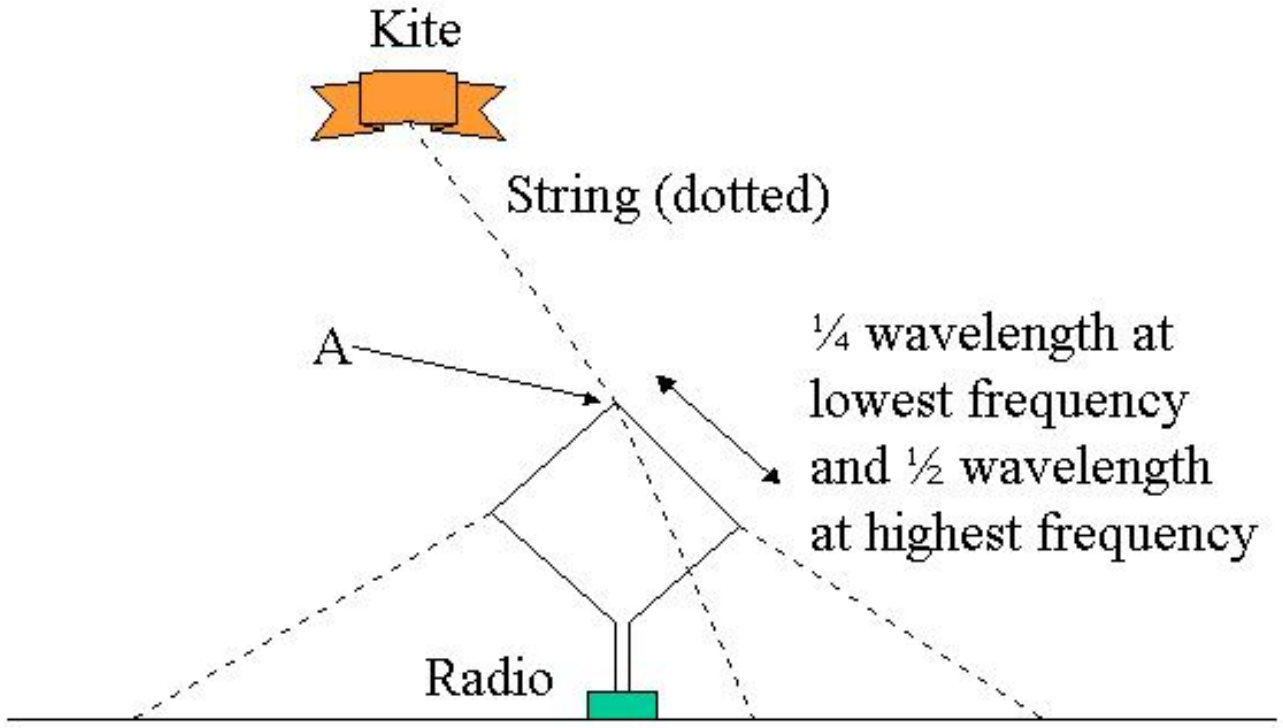
wavelength vertical seems to have been the best performer.

Some further thoughts are to use an end-fed half-wave inverted vee configuration with the feed point at ground level and the current point at the peak of the vee. This would be a good choice for near vertical incidence communications on the lower band and DX on the higher band (say 40m and 20m).



I did a test run of the configuration above on 4th July 2001. The aerials seems to work adequately. The kite needs to be well above the apex of the aerial to be clear of the vortices that exist around hill tops. Aim for 150 feet above the ground before the aerial is attached.

A further option would be a diamond aerial. This can be used on two frequencies (an octave apart) by simply opening or closing the circuit at point A. With a short at point A it would work on the lowest frequewncy (7MHz for example) while an open circuit would allow the aerial to work on 14MHz. This would be a useful scheme for K1 40/20 owners. The loop can be fed with open wire line.



Caution

Avoid power lines. Avoid aircraft (my kite has been up to over 1000 feet). Avoid lightening. Avoid static rain. Any or all of these could get someone killed.

Buying your kite

Jalbert Aerotechnology don't seem to be on the web so perhaps they have gone. There are however other Parafoil manufacturers about.

More general information about parafoil kites here.

Sources of kites on the WWW - **you** judge if they are ok!

1 2 3 4

[Home]

Kites for Lifting Antennas

The Weather Service both in the US and abroad, made extensive use of kites for lofting weather instruments in the period from 1880 to 1920. Also kites were tested for man lifting in the same period, for applications like military observer. What was accomplished was just remarkable.

Versions of the winged box kite was the champion in all cases. But that was before the unspared, parafoil kites we have today were invented. The standard was a box kite called the Blue Hill Meterological Box Kite. It had a rectangular section when viewed from the front.

Today, there are special kite stores, so one is not limited to what you can build from plans in books, or the cheap paper diamond kites sold in toy stores. Any good kite store will be loaded with the fancy, and expensive aerobatic kites, but they should also have excellent load lifters, if you know what to ask for and look in the dustier corners:

---

1. **The Classic Box Kite**. This was a standard, considered by some hard to fly. Multicell box kites are complex to build, but work well. Some of the variations include a tetrahydron shaped box kite, but it is mostly good looking, not a good lifter like the classic box kite.
2. **The Winged Box Kites**. Probably the best of the rigid kites. The triangular winged box kite is a great lifter. The French Military Kite, is a classic. Dihedral wings on box kites add to stability and lift. One of the most famous man lifting kites was a several-celled box kite with small wings, the Cody Manlifter. [No, that is not Buffalo Bill Cody, but a person often confused with him, Sam Cody, an early kite/aeroplane pioneer.]. You can even purchase modern versions of the Cody Manlifter in several small sizes. It is an excellent design.
3. **The Parafoil**. A commercial company makes four sizes usually available. The sizes are rated in square feet of area and go from pocket size to a bit less than four feet wide and a bit less than five feet from front to back. They are attractive because they fly at higher string angles than the box kites, more vertical. They pull like a deamon, with the largest size recommending 250 pound test flight chord. They wad up into a stuff sack that is very easy to transport, since they have no sticks or spars.
4. **The Sleds**. In its simpliest form the sled is a flat piece of cloth tied at the right and left side to the bridal, with two spars running from front to back to hold it rigid in one axis only.

The **parafoil** along with the **sleds** have a disadvantage in gusty winds. They will collapse in the air and fall like a rock, all the way to the ground if they do not reinflate in time. The **rigid kites** are far more immune to gusty winds. But they have spars etc, that make them harder to pack, and heavier. Some of the

kites of the 1890's had so many spars, cells, cross supports, braces, etc. that they were as complex as the early aircraft. The Cody Manlifter was very picky about having the fabric tight, thus having curved edges.

The Scott Sled and its variants is a good stable kite that is spared in one direction only, front to back. The Allison Sled is a variant of it. Various versions add vents in assorted patterns which are alleged to increase stability and tolerance for wind variations. For wind measurements I have flown simple tethered sleds in the trade winds for weeks at a time. Our record was nine days that a sled stayed up, staked to the ground, day and night, through land breeze/sea breeze transitions, with no intervention. It was finally collapsed and knocked down by a nasty gusty roller of air. It took seconds to shake it open and relaunch it.

Other kites that are useful as lifters include double triangular winged box kites, Coynes, and even the Malay kite which is the stock diamond shaped kite just about everyone bought at the toy store as a kid and played with. A Malay is certainly easy to build and will loft a small antenna wire, but it is unstable without a drag producing tail, at least when compared to things like winged box kites and parafoils.

One person has made a study of sewing a type of vented sled kite that is kind of a cross between a Scott Sled and a Parafoil. He has flown some very large ones which he has made and used them to do heavy lifting, and launch a squadron of parachuting stuffed bears.

The U.S. Weather Service standardized on the "French Military Kite", a triangular winged box kite. The kite is like a classic box kite, but only with three sides to the box instead of four. It also has dihedraled wings added to one surface which greatly increases its side to side stability. The Weather Service flew trains of these kites to increase lift and stability since they sampled air at multiple levels, much like space diversity receiving. One such train of kites was flown to 27,000 feet! The pull was so great that these trains of large kites were flow on piano wire, a very dangerous practice for the amateur. The Blue Hill Box Kite was another weather bureau standard. It was a large box kite with a rectangular cross section cells about three times as wide as they were high, but otherwise similar to the standard two cell box kite. It had one fore cell and one aft cell with fore/aft spars holding the two cells in alighnment.

On the top of Mauna Kea (13,380 feet) I have flown a small sled through two changes of wind direction and been able to look up and see the kite string make a full spiral above my head as it rose the several thousand feet to the kite itself. The string was very small, but Kevlar with an incredible breaking strength that would slice off fingers before it broke. One weather service report from the 1920's reports a train of kites that made nine revolutions of a spiral as it ascended above an observer through multiple changes of wind directions.

You should not use the antenna wire as "kite string", but support the kite with a flying string and parallel or hang the antenna structure from the flying string. A kite carrying away in a wind with a trailing antenna wire is a dangerous situation. You should make sure that if the kite breaks free, it cannot take the antenna wire with it. One way of helping to insure this, is to use a stouter cord below and through the

antenna bearing section, and a lighter cord above, so if the kite carries way, it is more likely to break off above the antenna section.

Big Kite + Stiff Trades = Major Tension

One last word of warning. If you get a decent large kite, like a six foot triangular box kite or a big size 15 Parafoil, get some serious kite cord, something well over 125 pounds. Also purchase some heavy duty leather gloves with wrist protection flanges on them, like gauntlets. Do not underestimate the potential dangers of flying a large kite in any kind of wind.

These big kites are renouned as lifters for a reason. A large kite needs preparation before launching into any wind regime. A modern ripstop nylon kite in a large size is NOT the toy paper dimestore kite of your youth, flown with cheap cotton twine.

A toy paper kite will tear up in too much wind. The flimsy cotton string used will break easily and is not prone to rope burns, slicing wrists to the bone or popping off fingers at the joints. The flimsy spars of the dimestore kite will snap easily if overstressed. Thus it is relatively safe to fly such a kite.

A kite made of ripstop nylon, carbon filament spars and flown with heavy duty nylon cord is a very different item that must be respected.

The first time you try to hang onto a large parafoil in a stiff 30 mph wind, you will know what I am talking about! Rope burns can happen quickly and cut to the bone or worse. Hand protection is mandatory. Rope burns are terrible. The skin and muscle are "erased" away leaving a deep trench that is slow to heal and very, very painful.

It is well worthwhile to invest in a small parafoil or winged box kite suitable for 40 or 50 pound line and test fly it to get a feel for heavy duty kites. Remember, a kite twice as large is going to have a pull four times as great or more. Also pull goes up with the cube of the wind speed!

The large military kites were flown from special trucks that carried large drum wenches. These kites were only about three times larger than kites easily available at kite stores today and routinely lofted observers to several hundred feet in WWI.

An interesting "target kite" was used in WWII that had a metal rudder on a 10 foot tall Malay type kite with a black aircraft silhouette painted on it. It was towed behind a jeep with an "operator" in the back who used a pair of control lines to manipulate the rudder on the kite. The jeep would go around an oval course while 50 caliber machine gun trainees would blast away at the kite above.

A friend had one of these and we would fly it in Oklahoma from the ground using a 4x4 inch oak wood beam, 6 feet long, as a flight handle. We would stand behind a telephone pole and let the pole take the

strain of the kite on the 4x4 handle and still be able to loop the kite and steer it by rocking the 4x4 back and forth behind the pole. It was impressive to see the divot ground into the 4x4 where it was bearing against the telephone pole.

One report of an experiment with 160 meters raised several interesting points. The objective was to fly a kite next to the ocean and use the Pacific Ocean as a ground plane. It worked extremely well but had two major problems.

1. It picked up literally volts of signal from the local high power AM broadcast stations, blotting out the front end of the receiver. Clearly, unless you are quite far from such a broadcast site or sites, you need to have a carefully built filter box. Probably bandpass with the major rolloff being on the low side of 1.8 Mhz. Possibly even a tunable notch section in the middle as well. Of course the usual static grounding of wind generated static electricity is essential. Any kite antenna MUST have a path to ground. A choke or a high value carbon resistor should work. In the case of a 160 meter antenna, the AM broadcast blocking filter could easily be designed to have as its first component something to provide a DC path to ground.

2. A kite that has fallen into the ocean is a monster sea anchor. Rip tide can make a kite impossible to retrieve. Kites have been built with collapsible harnesses. This is probably a great idea. Basically one examines the harness that holds the kite in the right aspect to the wind stream. Find a critical part that sets the angle of attack. Make this with a link of much weaker string than the rest of the harness and the flight line. Thus a abusively strong yank on the flight chord will break the harness causing the kite to flatten out and stream. Such a harness modification would have made it possible to retrieve the kite in the ocean. Since fancy lifting kites can cost a hundred dollars of more, getting the kite back is a priority.

Finally, consider a 1/4 scale model designed for 40 meters. It will solve a lot of problems and give a lot of experience at much less expense and hazard. Anything for 160 meters is just plain BIG.

One last word, if the kite is big or flow high, you can get into trouble with the Federal Aviation Administration. Here is a sample peek at typical FAA regulations. For most up to date information you should check the FAA's own web pages:

FAA Part 101

Return to UH Ham Club Home Page

06/02

W2IK's LADDER-LINE BACK PACK SPECIAL

GREAT FOR CAMPING OR EMERGENCY COMMUNICATIONS

BOB HEJL - W2IK

1568 HITOMETER

EVER WISH YOU HAD A VERY SIMPLE, LIGHT WEIGHT HAM RADIO ANTENNA THAT COULD BE STUFFED IN YOUR BACKPACK AND DOES NOT NEED A TUNER? HERE IS A SIMPLE DESIGN THAT I BUILT MANY YEARS AGO AND FOUND IT TO BE VERY USEFUL WHILE CAMPING OR FOR EMERGENCY COMMUNICATIONS. IT'S GREAT IF YOU DON'T HAVE THE ABILITY TO TAKE ALONG MY "IK-STIC 2" ANTENNA (TO BUILD THAT ANTENNA, CLICK ON THE RED "IK-STIC 2" TITLE)

WHEN YOU ARE CAMPING, YOU HAVE TO MAKE COMPROMISES. THE SAME IS TRUE ABOUT CAMPING WITH HAM RADIO. USUALLY YOU'LL BE OPERATING ON 20 METERS IN THE DAY AND 40 METERS AT NIGHT. THAT BEING SAID, I DEVISED A VERY SIMPLE ANTENNA IN A SINGLE FORM THAT WILL COVER BOTH BANDS WITHOUT THE NEED FOR A TUNER.

TO BUILD THIS ANTENNA YOU WILL NEED

50 FEET OF 450 - 600 OHM LADDER LINE (NOT TWIN LEAD)
2 THREE INCH PIECES OF 1/2 INCH PVC TUBING
ONE CENTER CONNECTOR (DIPOLE TYPE-WIRE TO SO-239 CONNECTOR)

UNROLL THE COIL OF LADDER LINE AND CUT THE LADDER LINE IN HALF SO YOU HAVE TWO LENGTHS EXACTLY 25 FEET LONG. EACH LENGTH BECOMES A SIDE OF YOUR DIPOLE SYSTEM. AT THE EVERY END OF EACH WIRE, STRIP OFF THE INSULATION EXPOSING ABOUT 4 INCHES OF BARE WIRE. ON ONE SIDE OF EACH LADDER LINE TWIST THE BARE WIRES TOGETHER AND SOLDER ONE END OF EACH LADDER LINE TO THE WIRES ON THE CENTER CONNECTOR.
(SEE PICTURE BELOW)

AT THE OTHER END OF EACH OF THE LENGTHS OF LADDER LINE, TAKE ONE PIECE OF 1/2 INCH PVC TUBE AND SLIDE THE WIRES THROUGH 1/4 INCH HOLE DRILLED NEAR ONE END OF THE PVC TUBE. PULL THE PVC AROUND SO YOU CAN CONNECT THE WIRES TO EACH OTHER AND SOLDER THEM TOGETHER.
(SEE PHOTO OF THE END INSULATOR BELOW)
DRILL ANOTHER 1/4 INCH HOLE ON THE OTHER END OF THE PVC TUBE SO YOU CAN HANG THE DIPOLE SYSTEM UP USING NYLON CORD.

THE SPECIAL CUT

NOW JUST ONE MORE THING TO DO. IN ORDER TO MAKE THIS A 40 AND 20 METER ANTENNA, YOU NEED TO MEASURE 16.4 FEET FROM THE MIDDLE OF THE CENTER CONNECTOR AND ON ONE WIRE OF EACH SIDE OF THE LADDER LINE YOU WILL CLIP OUT A 1/4 INCH SECTION OF WIRE. IF POSSIBLE, CLIP THE SECTION OUT WHERE THERE IS A SEPARATING PIECE OF THE LADDER LINE WEBBING SO THE ANTENNA WILL BE PHYSICALLY MORE STABLE.

THIS WILL LEAVE YOU WITH TWO ANTENNAS IN ONE SYSTEM. ONE ANTENNA'S TOTAL DIPOLE LENGTH IS 32.8 FEET (16.4 FEET ON EACH DIPOLE LEG) OR A 20 METER DIPOLE, AND THE OTHER IS THE 40 METER DIPOLE AS THE WIRE IN LADDER LINE YOU DIDN'T CLIP WILL NOW RUN THE ENTIRE LENGTH OF EACH SIDE, CONTINUE AROUND THE END INSULATORS AND CONTINUE IN A HAIRPIN RUN AROUND TO THE OTHER (LOWER) SIDE OF THE LADDER LINE ON THAT SAME SIDE MAKING A LENGTH OF ABOUT 66.8 FEET TOTAL OR 33.4 FEET ON EACH LEG (LESS THAN THE MATH OF JUST ADDING AND SUBTRACTING BECAUSE YOU LOST LENGTH WHEN YOU SOLDERED THE WIRES THAT ARE IN THE END CONNECTORS) THIS IS FINE FOR A 40 METER DIPOLE.

THERE IS JUST A LITTLE INTERACTION WITH THE TWO ANTENNAS SO CLOSE TOGETHER. IF YOU WISH TO BE REALLY FUSSY, YOU MAY WANT TO DO THE WIRE CLIP ACTION AND CHECK THE TUNING USING AN ANTENNA ANALYZER. IF THIS IS THE CASE, DO NOT CLIP OUT A 1/4 INCH PIECE YET, JUST CLIP THE WIRE, MEASURE THE SWR WITH THE ANTENNA HUNG AND MAKE ANY CORRECTIONS (DO THIS BY RESOLDERING THE CUT YOU MADE AND MAKE ANOTHER CUT DEPENDING UPON THE ANTENNA'S RESONANT FREQUENCY). ALTHOUGH THIS ANTENNA IS NOT AS EFFICIENT ON 40 METERS AS A DIPOLE STRUNG OUT INSTEAD OF FOLDED, I HAVE USED THIS ANTENNA ON MANY OUTINGS WITH GREAT RESULTS AND IT IS MORE CONVIENT TO PUT UP OR STORE. YOU CAN EASILY COIL IT UP AND STORE IN YOUR BACK PACK. IT CAN BE PUT UP AS A DIPOLE, SLOPER OR INVERTED "V". MAKE SURE YOU HAVE NON-CONDUCTING CORD TO HANG IT UP. AND BY THE WAY, SINCE THE 40 METER SECTION LOOPS AROUND, ALL YOU NEED IS 52 FEET BETWEEN TREES TO HANG THIS ANTENNA AS A DIPOLE!!!

LADDER LINE ANTENNA COILED UP FOR STORING IN A BACK PACK SO IT CAN BE USED WHILE HIKING OR CAMPING

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HAM RADIO AND THE WORLD TRADE CENTER DISASTER
THE AFTERMATH OF WTC 9/11 - MY LIFE IN A SHELTER
OVERLOOKED ANTENNA CONSTRUCTION POINTS

HAM RADIO CAN BE HAZARDOUS TO YOUR HEALTH - NEW

"FAMILY PREPAREDNESS" -NEW

"IS YOUR COMMUNICATIONS GROUP VULNERABLE?"

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Not Everything Good is New: The Expanded Lazy-H

L. B. Cebik, W4RNL

Very effective antennas do not have to be exceptionally large or expensive. The latest designs and construction methods have their advantages--and also their costs. They tend to obscure some older designs of high merit as we forget to remember them.

Rotatable antennas are very effective, but for those unwilling or unable to put a tower, rotator, and sizable aluminum structure in the air, fixed position wire arrays can provide excellent gain. Most designs are bi-directional, but the side rejection is often sufficient to eliminate most QRM. If we have the trees or the poles to support the ends, and if we take the trouble to align the antenna in the most favorable directions for our intended operation, a wire array can work wonders. For example, a great circle drawn through my QTH in Tennessee with one end in VK-ZL land will have its other end in Europe. A bi-directional array might be just the ticket for much of my operating.

Many broadside arrays are flat-tops--that is, they require at least two wires with considerable horizontal space between them. For most purposes, I would need 4 supports. However, a design that has been around since wire became popular is the Lazy-H, a vertical stack of two wires fed in phase. The standard Lazy-H consisted of two 1 wl wires spaced 1/2 wl apart and elevated so that the lower wire was 1/2 wl above ground.

John Schultz, W2EEY, wrote in the November, 1968, *CQ* of the "Expanded Lazy-H Antenna." Bill Orr, W6SAI, recalled this antenna in one of his many columns during the 1980s. Another 15 years has gone by, so let's recall this effective array one more time. The key to expanding the Lazy-H is to increase both the horizontal and vertical dimensions by just a little bit.

If we increase the wire lengths from 1 wl to 1.25 wl, we have stacked extended double Zepps in our Lazy-H. The effect is to give us a bit more gain per wire and a significant amount more from the pair. Then, if we increase the spacing from 1/2 wl to 5/8 wl, we achieve approximately the maximum stacked gain possible with two simple wires.

Now, let's build one of these expanded Lazy-Hs for 10 meters.

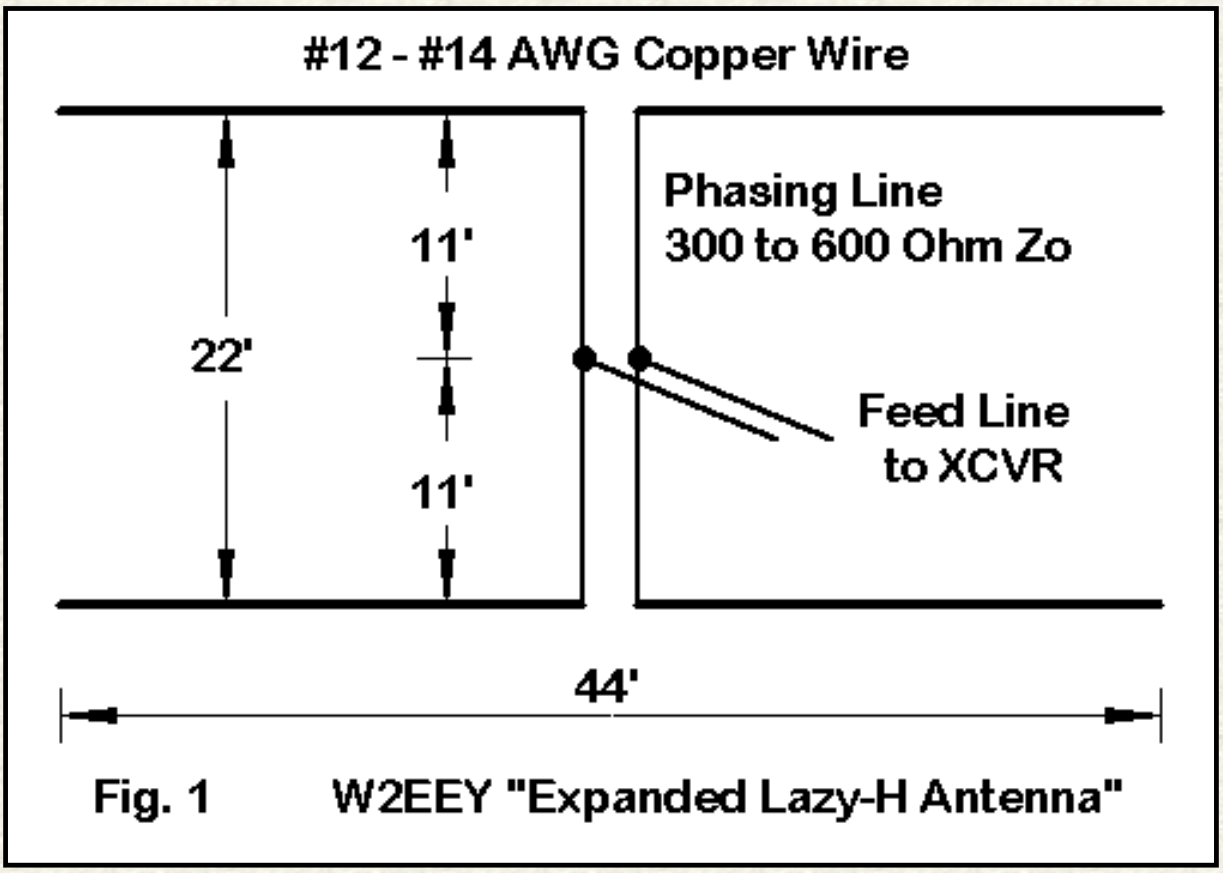

Fig. 1 W2EEY "Expanded Lazy-H Antenna"

Fig. 1 shows the antenna outline. For 10 meters, a length of 44' per wire is satisfactory and not critical: 40' to 50' will work, but the pattern on 10 meter begins to split up as we lengthen the antenna too far beyond 1.25 WL. Vertical spacing between the two wires need not be too fussy, but the recommended 22' gives us not only 5/8 wl at 10 meters but a usable spacing at other frequencies.

The recommended minimum of 1/2 wl at 10 meters is a bit low for optimal performance. I would recommend that a lower-wire height of about 44' be used, which places the top wire at 66' up. Lower heights will reduce the gain and elevate the TO angle from the figures I shall present as we think about this simple array.

The mechanical beauty of the Lazy-H design is that it requires only two supports--although fairly tall ones. The electrical beauty of the antenna is that it provides excellent bi-directional performance from 10 meters through at least 17 meters, with good performance down to 30 meters. It can also be pressed into service on 40 meters without much difficulty.

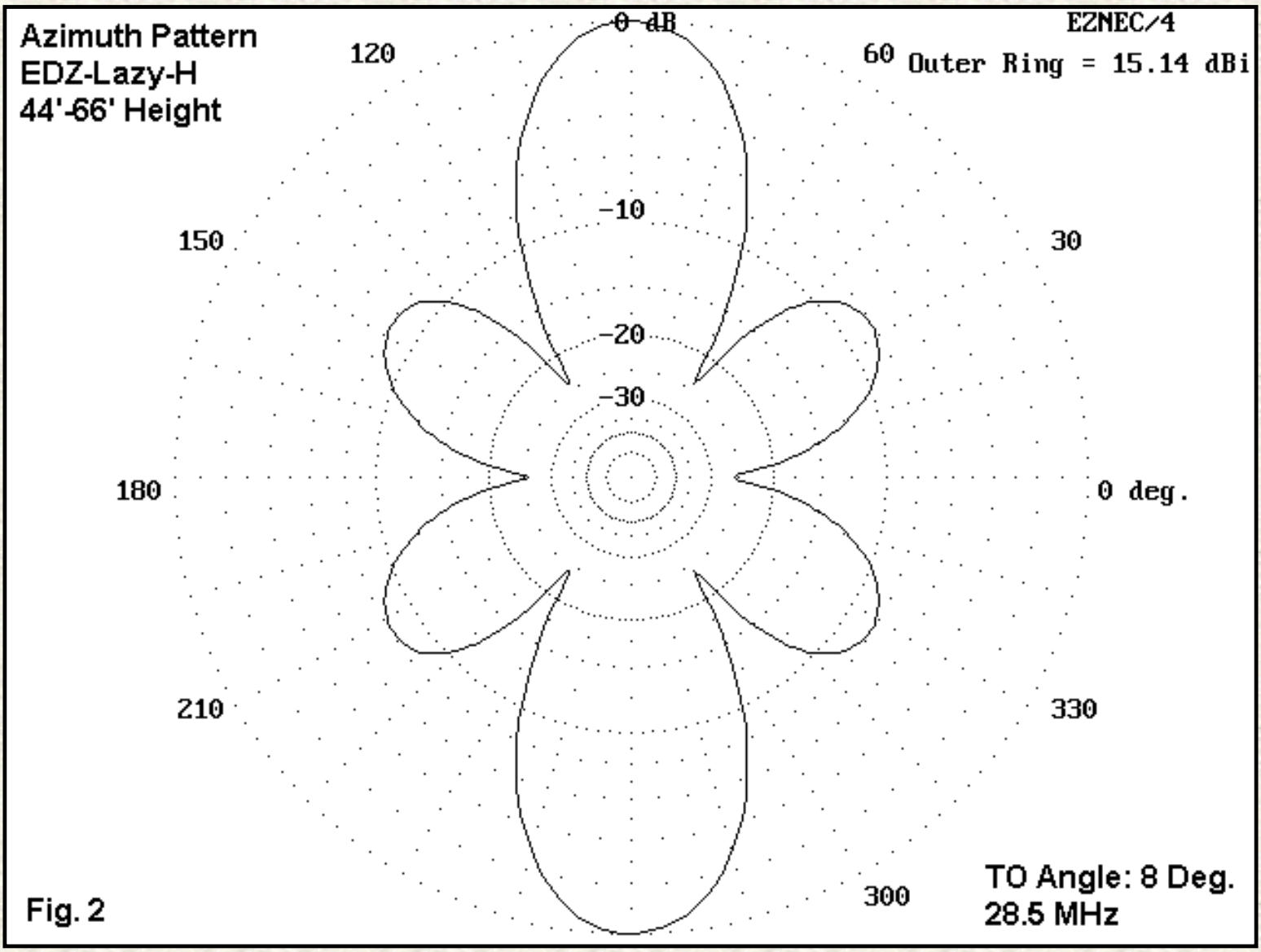

Fig. 2 shows the azimuth pattern of the antenna on 10 meters at an elevation angle of 8 degrees. The phase-fed array still retains the EDZ "ears." These ears are the beginnings of the multi-lobe pattern that emerges as the antenna wire length grows toward the 1.5 wl mark.

On all bands below 10 meters, the length of the antenna is under the EDZ mark, so the pattern is bi-directional with single lobes each way. In fact, at 15 meters, the antenna becomes a standard Lazy-H: two 1 wl wires spaced 1/2 wl apart vertically.

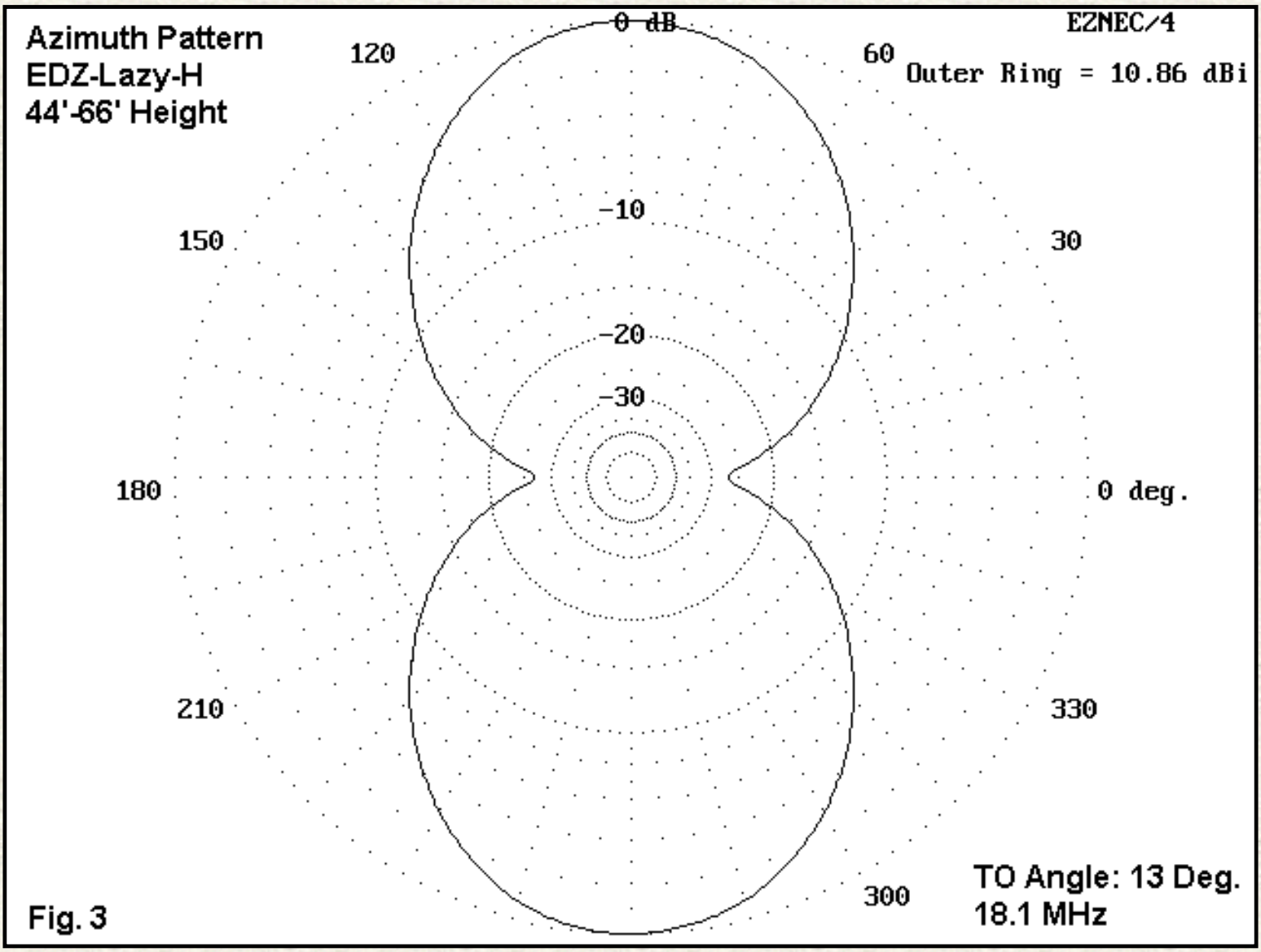

Fig. 3

In **Fig. 3**, we see the 17-meter pattern at its elevation angle of maximum radiation of 13 degrees. As the tables below will show, the lobes become wider as we reduce frequency and narrower as we increase frequency.

For a more systematic view of anticipated performance on all of the possible bands on which we might use this one Lazy-H, here is a table of modeled performance over average ground, with the lower wire 44' up. The table lists the usual gain and TO angle data, but also adds numbers for the vertical and horizontal beamwidths between the -3 dB (half-power) points. This date is useful in determining the azimuth coverage of the antenna in each direction and in estimating the elevation angles to catch the skip for varying circumstances.

Freq MHz	Max. Gain dBi	TO angle degrees	Vert. BW degrees	Hor. BW degrees
28.5	15.1	8	9	31
24.9	14.6	10	11	41
21.2	12.5	11	12	52
18.1	10.9	13	14	61

14.15	9.0	17	18	73
10.1	8.1	24	27	85
7.15	6.4	33	44	99

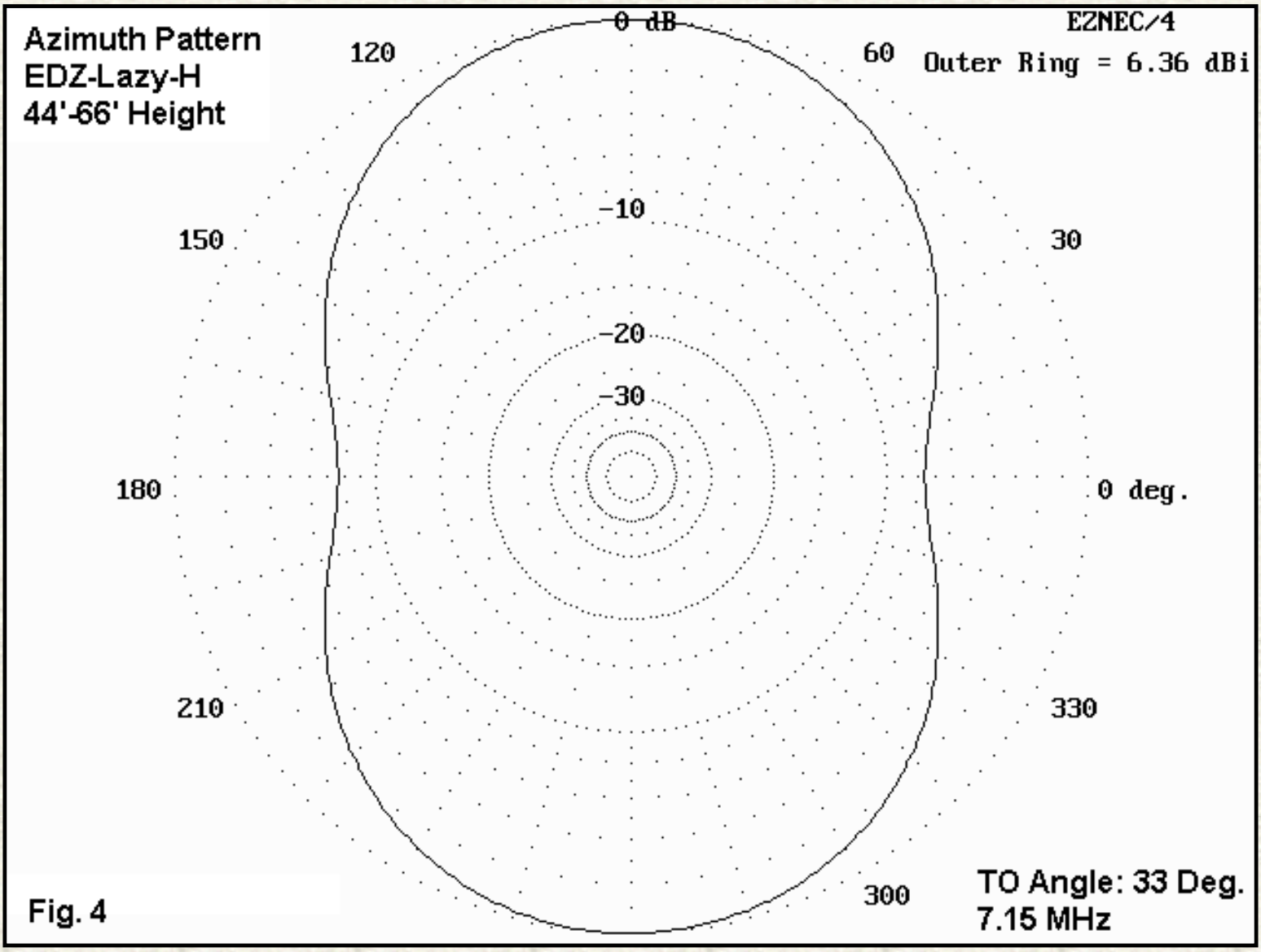

Fig. 4

By the time the antenna's operating frequency is lowered to 40 meters, the pattern becomes a broad oval with a fairly high TO angle, as shown in **Fig. 4**. However, sufficient radiation occurs at lower angles to make it usable for general purpose communications on that band.

How good is the antenna's performance? I could use any number of comparators here, but the simplest would be a single 44' wire placed 66' up in height, the same height as the top wire of the array. The usefulness of this comparison is that it helps reveal something of the array's characteristics.

Freq MHz	Max. Gain dBi	TO angle degrees	Feedpoint Z R +/- jX Ohms
28.5	10.5	7	150 - j 695
24.9	10.4	8	620 - j1700
21.2	9.0	10	4200 + j 850

18.1	8.6	12	835 + j1560
14.15	7.7	15	190 + j 490
10.1	7.6	20*	56 - j 105
7.15	7.0*	29*	24 - j 600

The starred gain entry for 40 meters indicates that the single wire at this frequency shows more gain than the array (by about 0.6 dB). In the TO angle column, the starred entries indicate that the single wire shows a significantly lower angle than shown for the array. Both phenomena are related. The array elevation angle of maximum radiation is a composite from radiation from both wires, with the lower wire radiation raising the angle of the final composite pattern. The difference is slight until the very lowest bands on which we might press this antenna into service. On 40 meters, the lower wire is just over 1/4 wl above ground, so that it raises the overall pattern angle of the array by a goodly amount and provides slightly less gain than the single wire that is about 1/2 wl up.

From 20 meters on up, the Expanded Lazy-H shows good gain over a single wire. The benefits increase the higher one goes in frequency, up to the break-up of the pattern when the wires are longer than 1.25 wl. It is certainly possible to scale the antenna for maximum benefits at a lower frequency, but that lower frequency of maximum gain will become the highest frequency at which one can use the array and still have a bi-directional pattern with a single main lobe off each side of the wires.

The single 44' wire also shows a wide variation in feedpoint impedance according to the length of the wire. The 10-meter value is typical for an EDZ. The 15-meter value is also typical, but of a 1 wl center-fed wire. Parallel feeders and a highly competent antenna tuner would be needed for this antenna. However, careful analysis of the impedance excursions along the chosen feedline can minimize the chances that the tuner antenna terminals will see either a resistance or a reactance value outside its range of adjustment.

So far, I have given no figures for the feedpoint impedance of the Expanded Lazy-H. The two elements in Fig. 1 are fed in phase by the simple expedient of using equal lengths (11') of line to a center point to which we attach the parallel feeders going to the antenna tuner. There are two controllable variables that will affect the feedpoint impedance at the junction with the main line to the shack. One is the length of the lines, which we have set at 11' each. The other is the characteristic impedance and velocity factor of the phasing lines. I shall not here explore other phasing line lengths, but instead shall show some anticipated feedpoint impedances for each band using three different phasing lines. One will be a 450-Ohm, 0.95 VF line, typical of windowed vinyl-covered lines. Another will be 300-Ohm, 0.8 VF line, typical of good quality TV line. The third will be 600-Ohm, 1.0 VF line, which might be bought or built from wire and spacers.

	Feedpoint Impedance (R +/- jX Ohms)		
Freq	450-Ohm	300-Ohm	600-Ohm
MHz	0.95 VF	0.8 VF	1.0 VF
28.5	65 + j 425	115 + j 570	105 + j 610
24.9	17 + j 115*	11 + j 140*	30 + j 140
21.2	22 - j 15*	10 + j 38*	40 - j 50
18.1	45 - j 125	16 - j 26*	90 - j 230
14.15	385 - j 395	75 - j 150	1050 - j 350
10.1	50 + j 105	40 + j 65	50 + j 155
7.15	10 - j 95*	6 - j 80*	13 - j 90*

Starred entries represent very low resistive components to the feedpoint impedance which might present larger excursions along whatever line is chosen as the main feedline to the shack. Note that the starred entries are fewest with the 600-Ohm phasing line. Once more, it is worth noting that these numbers are derived for general guidance from models. Variations will emerge from the actual construction of the antenna and from conditions and clutter at the antenna site.

One question that almost always emerges with respect to comparing the single wire and the array gain figures for 10 meters is this: how can the array deliver over 4.5 dB gain over the single wire? The answer is straightforward if we compare elevation patterns for the two antennas. **Fig. 5** tells the tale.

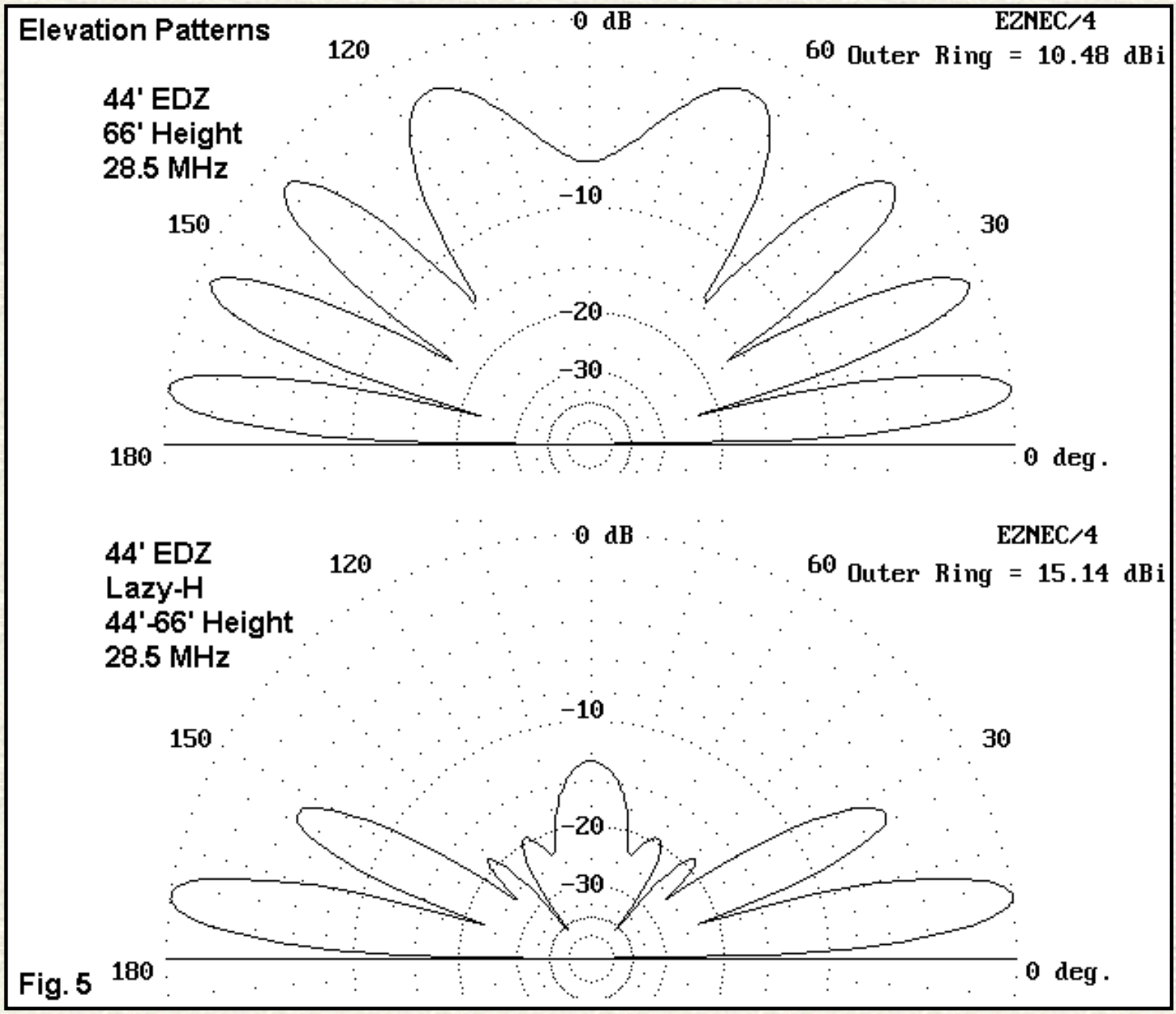

Fig. 5

Like any single-wire antenna, the EDZ at 66' on 10 meters shows an array of nearly equal-strength vertical lobes: 4 to be exact. In contrast, the upper lobes of the Expanded Lazy-H are suppressed leaving a single dominant lobe and a secondary lobe well over 4 dB weaker. All other lobes are down by 12 dB or more.

The array tends to waste far less power at very high angles of radiation compared to the single wire. This comparative pattern, with variations, tends to hold true down through 20 meters.

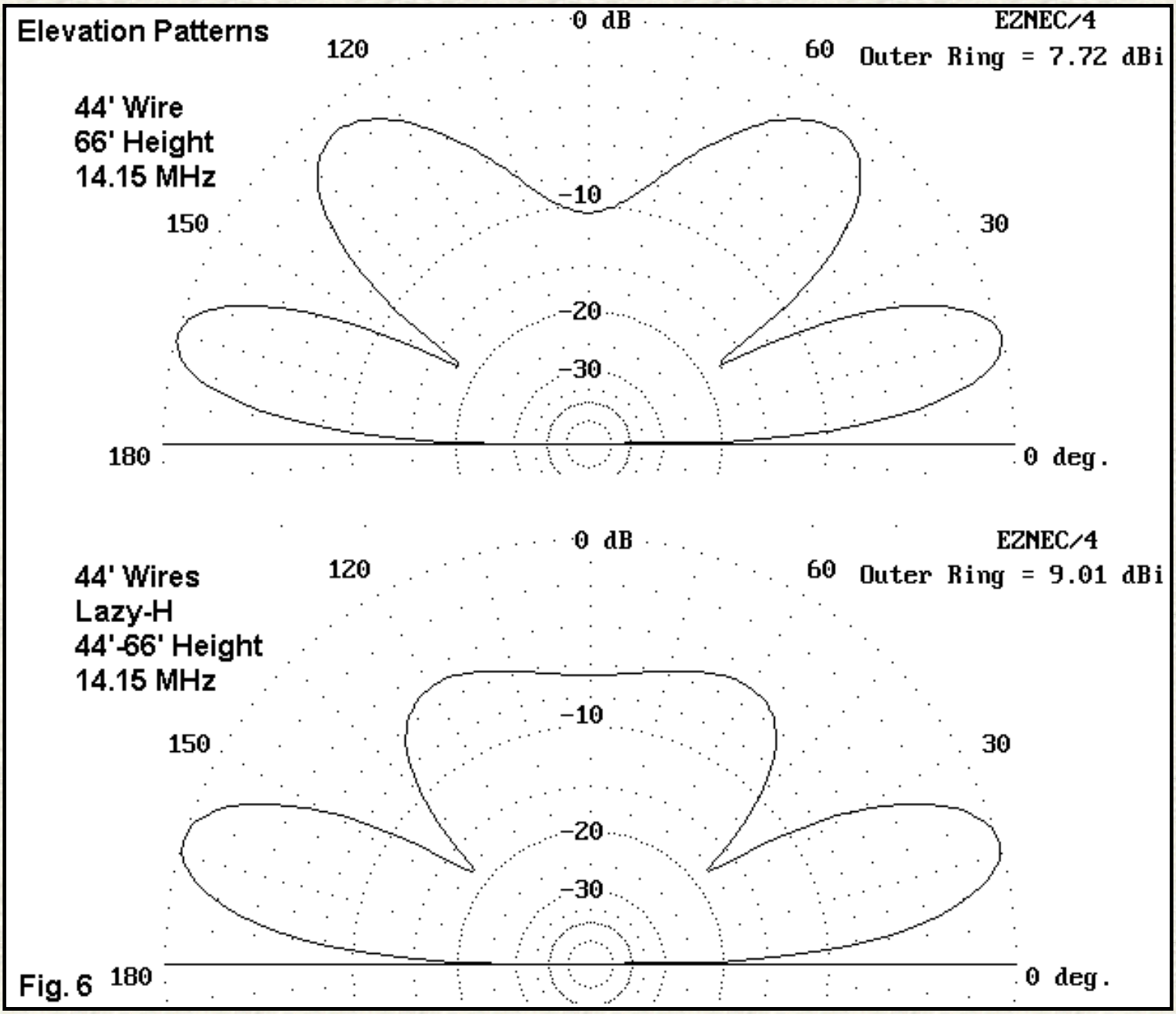

Fig. 6

On 20, the effect is less pronounced but still easily measured, as shown in **Fig. 6**. The area enclosed by the upper lobes of the single wire at the top of the figure is distinctly greater by a considerable margin than the area enclosed by the upper lobe (barely discernable as a double lobe) of the array. The difference in area (assuming that the azimuth patterns are comparable, as they happen to be in this case) is a rough measure of the added power appearing in the lower lobes. In this case, that additional power shows up not only in the maximum gain, but as well in the vertical beamwidth. The phased feeding of vertically stacked horizontal wires has benefits hard to match in a typical flat-top wire array.

Along side the benefits come some limitations. The Lazy-H requires a pair of tall supports and is suited to the antenna farm with more tall trees than money. It is possible to lay out more than one of these

inexpensive antennas in order to cover additional regions along the horizon. It is likely that no special treatment will be needed to detune unused arrays to prevent them from altering the pattern of the array in use. Either leaving the shack end of the unused feedline open or shorting it will introduce to the wire feedpoints sufficient reactance to detune the wires. However, this is a facet of multiple array installation that the builder should keep in mind. Sometimes Murphy dictates that nothing will work to prevent interaction short of greater physical separation of the arrays.

The Expanded Lazy-H is an outstanding bi-directional array for 10 meters in the design given here. Its performance holds up well down through 20 meters, and we can press it into service on lower bands. It takes up very little room horizontally in the yard, although a couple of optimally spaced tall trees certainly can aid the installation process. The wires for the elements and the phasing lines, as well as the feedline to the shack and the UV-resistant support ropes, are certainly inexpensive compared to the cost of a tower, rotator, coax, and commercial aluminum antenna. It is a design worth recollecting every 15 years or so just to make sure that we do not forget it.

Can the extended Lazy-H be converted into a directional beam for some or all of the bands that it covers? To find out--at least in principle--see **"Curtains for the Extended Lazy-H."**

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In Brief

March 22, 2001

(In alphabetical order:)

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ARRL Products: Pins, Patches, Maps, Gift Ideas & Supplies...
(More)

MINILOG -- Pocket-size logbook.

QST Binder--Large -- 8.5" × 11" for *QST* 1976 and after.

Life Membership Patch - round -- Round embroidered patch, 3-inches. Red, white, black and gold threads.

Flag Patch -- Cloth patch with ARRL flag design.

H. P. Maxim Mug (photo) -- Ceramic coffee mug (10 oz) with photo of ARRL Founding President, Hiram Percy Maxim, W1AW.

AMSAT-NA 2001 Annual Symposium issues call for papers: The first call for papers has been issued for the AMSAT-NA 2001 Annual Symposium, set for October 5-6 in Atlanta, Georgia. Papers may be presented by the author during the symposium or simply offered for inclusion in the *Symposium Proceedings*. Subject matter should be of general interest to Amateur Radio operators involved in satellite communications. Suggested topics include operating

techniques, antenna design and construction, spacecraft design and construction, current mission status, proposed satellite missions, telemetry acquisition and relay, satellite microwave projects, etc. An outline abstract of the proposed paper should be submitted as soon as possible. The final date for abstracts is June 30, 2001. Completed papers must be received no later than August 15. Electronic submittal is preferred. The format must be either *MS Word*-compatible or plain text. For security purposes, Symposium Chairman Steve Diggs, W4EPI, asks authors to condense the document file and send it as an e-mail attachment to w4epi@amsat.org.--*AMSAT News Service*

Array Solutions acquires RF Applications marketing rights: Array Solutions has acquired the exclusive marketing rights to RF Applications VFD, P-3000 and P-5000 wattmeters and the IBS-1 Intelligent Band Switch. RF Applications will continue to manufacture and support these products and will honor all existing warranties. "We are excited that Array Solutions has elected to add our products to their product line," said RF Applications President Bruce R. Knox, W8GN. "We feel that our products need better exposure to the marketplace, and Array Solutions' market presence is a perfect answer to that need." Array Solutions President Jay Terleski, WX0B, said the RF Applications' power-measurment devices will be an important part of the company's vertical array-phasing products it's introducing this year. RF Applications has announced that it no longer will manufacture its P-2000A, P-5000A and P-2000CW wattmeters. The WinWatt will continue to be marketed and sold exclusively by RF Applications.

California ARES team gets into a stink--and the nose knows! (This is not a story for the squeamish, so if you're of a sensitive nature or just ate, perhaps you should just skip to the next item--*Ed.*) Amateur Radio Emergency Service members in Ventura County, California, recently got into a stink--quite literally. As Ventura County ARES District Emergency Coordinator and RACES Radio Officer Dave Gilmore, AA6VH, explains, this story may sound like just so much horse manure--and you'd be right. Three and a quarter acres of it, 15 feet deep, and all on fire. Manure from an area race track is mixed with hay and used to compost mushrooms. The compost piles have been known to spontaneously combust, and that's what happened recently. "The smoke and smell even traveled to Los Angeles County to our west and south, and to Santa Barbara County to our north, Gilmore said. Fire officials asked Ventura County ARES/RACES to help determine the direction of the smoke cloud. "Unfortunately, it was foggy and hazy, making it hard to determine where the smoke ended and the haze began," Gilmore said. "The only way to tell was by sniffing the air." Hams were sent out to sniff various areas and report back. Gilmore says that those on "sniff patrol" periodically were forced to refresh their olfactory capabilities. "After a couple of whiffs, most people get used to even foul odors and can mistakenly believe the odors have gone when they actually have not," Gilmore explained. In addition to fire officials in two jurisdictions, air pollution and health officials got into the act. For a variety of very good reasons, it was decided the best way to put out the fire was to cart off the burning manure to a nearby field to disperse and extinguish. "Unfortunately the dispersal and movement of the fuel would cause a temporary increase in the amount of smoke produced by the fire," Gilmore said. By the second and third days, ARES/RACES members were stationed to report when the smoke and smell became intolerable, so that operations could temporarily halt until the smoke level decreased. After the weather cleared, one operator was stationed atop a nearby hill to observe the smoke trail and direct the operators around as the smoke plume shifted. By the third day, Gilmore said, enough of the compost had been moved to decrease the smoke level to the point where it no longer posed a problem.--*Dave Gilmore, AA6VH*

Crowd welcomes HRO-Anaheim back to original location: A huge crowd was on hand to

HRO Vice President Robert (left) and President (and Robert's father) Bob Ferrero, W6KR and W6RJ, flank HRO-Anaheim Manager Janet Margelli, KL7MF, as she cuts the ceremonial ribbon to signify the return of HRO to its original location at 933 North Euclid in Anaheim. [K7JA Photo]

celebrate the return of Ham Radio Outlet to its original location at 933 North Euclid in Anaheim, California. That location was destroyed by fire in January 2000 and the store has been in nearby temporary quarters since then. Representatives from Alinco, ARRL, ICOM, Kenwood, NCG, Pryme, RF Parts Co, Sangean, US Tower, and Yaesu were busy showing off their latest wares, Gordon West Radio School handed out Amateur Radio leaflets to passers by. HRO-Anaheim is the largest Amateur Radio store in the Los Angeles area. Ham Radio Outlet has 12 stores throughout the US, five in California. "It was great to welcome so many of our local customers to our Grand Re-Opening event," said HRO-Anaheim Manager Janet Margelli, KL7MF. "Their support through the long and arduous rebuilding process has been greatly appreciated by all of us in the HRO organization."

Dayton Contest Dinner tickets on sale: The North Coast Contesters have announced that tickets for the ninth annual Dayton Contest Dinner now are on sale. Sponsored by *CQ* magazine, the master of ceremonies will be *CQ* Contest Editor John Dorr, K1AR. Featured speaker will be WRTC-2000 Chairman Tine Brajnik, S50A. The names of 2001 Contest Hall of Fame inductees will be announced by the *CQ Contest* Editor Bob Cox, K3EST. The dinner will be held Saturday, May 19, 6:30 PM, in the Van Cleve Ballroom, Crowne Plaza Hotel, Fifth and Jefferson streets (next to the Convention Center), in downtown Dayton. Tickets are $30. Seating is limited to 300, no reserved seats. For tickets and additional information, contact Craig Clark, W1JCC, Radio Bookstore, PO Box 209, Rindge, NH 03461; Place orders by telephone weekdays 10 AM to 6 PM Eastern, 800-457-7373 (toll-free) or 603-899-6957, or by FAX 603-899-6826 anytime, or via e-mail to nx1g@top.monad.net. Credit cards are accepted but no CODs. E-mail orders must include charge card, call sign and return address information. Tickets will be sent via first-class US Mail no later than May 8. Deadline for ticket orders is May 7, 2001.

EMCOMM 2001 is March 31: The second annual Emergency Communications Conference will be held Saturday, March 31, 2001, at Bishop Quinn High School in Palo Cedro, near Redding, California. The all-day event begins at 9 AM; registration gets under way at 8 AM. EMCOMM 2001 will include seminars and workshops on topics such as search-and-rescue radio communications, emergency antennas for VHF and HF, choosing batteries and emergency power sources, public relations and working with the media, an overview of the incident command system and SEMS, "mutual aid" practices, and handling formal traffic. The National Weather Service will offer a presentation on SKYWARN. EMCOMM 2001 is sponsored jointly by ARRL Sacramento Valley Section Amateur Radio Emergency Service and the California Office of Emergency Services Auxiliary Communications Service. Anyone interested in volunteer emergency communications is welcome. For more information, visit the Sojourners Net Web site, http://www.qsl.net/k6soj or write EMCOMM 2001, c/o D.W. Thorne, K6SOJ, PO Box 16, Macdoel, CA 96058.--*Donna Ferguson, N6SVV*

FCC chairman touts "enforcement model": Speaking at the CTIA 2001 Wireless show in Las Vegas, Nevada, this week, FCC Chairman Michael Powell said the FCC is rethinking its business model. According to an *Infoworld.com* report Powell told the gathering that the FCC is "reviewing the optimal organization and structure" with an eye toward a more-responsive FCC. "We are putting increasing emphasis on an enforcement model as opposed to a regulatory model," he said, in order to speed up the FCC's decision-making process. "When you cheat, we'll get you at the back end." Powell also said there's a delicate balance for agencies such as the FCC between fostering innovation but not imposing it.

First QSO reported between ionic fluid antennas: The Live-Wire Group has recorded the first liquid antenna-to-liquid antenna contact on Saint Patrick's Day, March 17--and that's no blarney! The Live-Wire Group currently is experimenting with the liquid antenna concept. Participating in the 17-meter SSB contact were WH2AAT in Orange Park, Florida, and N9ZRT in Green Bay, Wisconsin. Both stations were using 10-foot tall by two-inch wide "columns of ionic fluid" (in this case, concentrated saltwater). Also participating in the QSO was W8ZU in Glen Rock, Pennsylvania, who was using a conventional antenna. Members of the Live-Wire Group have been experimenting with the Ionic Fluid Antennas (IFAs) for more than six months, and they report excellent results on the antenna's performance. In most cases the liquid antennas are operated in the vertical position. The RF is fed into the base of the antenna through several three-inch long copper probes that are exposed to the conductive liquid. Live-Wire members continue to experiment with this antenna concept in various forms including liquid dipoles and "pumpable-to-resonant-length" verticals. For more information, contact the Live-Wire Group, http://www.wireservices.com/livewire.html.

The 10-foot tall, two-inch diameter IFA antenna on N9ZRT's jeep. [N9ZRT Photo]

M0BMU wolfed in Massachusetts: There's more transatlantic news on the LF front. John Andrews, W1TAG, has successfully copied a WOLF digital signal from Jim Moritz, M0BMU, in the UK. "WOLF has successfully crossed the pond," Andrews exulted in an e-mail to the LF reflector after only a day of listening. Andrews reports receiving "two lines of clean copy" between 0135 and 0200 UTC on March 19 just below 137.5 kHz. Moritz also was pleased. "Many thanks to John for this report and his perseverance," he said. "Now QRSS is not the only way to get across the pond on LF!" QRSS is very slow-speed CW that's been used for LF contacts, while WOLF is a robust digital mode. Moritz reports using a Racal-Decca 5501 Class D power amplifier (1200 W PEP output) with a Racal 9084 signal generator as a synthesizer plus homebrew phase-keying, waveform-generating and amplitude modulator circuits. His antenna is a 40-meter long inverted Vee, 16 meters high in the center and 9 meters off the ground at the ends. He estimated his ERP at 1 W. "I also received a report from ON7YD, who was able to obtain perfect copy on my signal from a short recording," said Moritz, who plans to run additional tests on 137.500 kHz.

National Weather Service gets a MARS station: For unstinting public service by Amateur Radio

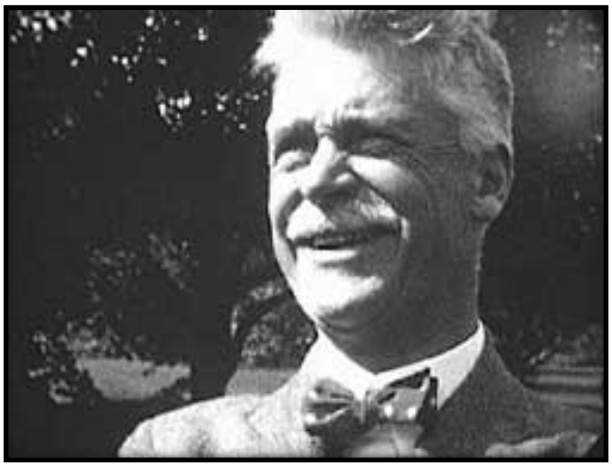
there's no better example than the SKYWARN partnership between hams and the National Weather Service. But even that wasn't enough during the 1998 ice storm that knocked out telephone communications across large areas of the Northeastern US for up to several weeks. During that emergency the NWS forecast office in Taunton, Massachusetts, outside Boston, lost contact with the NWS office in Gray, Maine, near Portland. Although dozens of SKYWARN members were activated throughout the long outage, their normal mode is VHF, and Gray is way out of VHF range of Taunton. Thus Taunton was unable to access Gray's vital weather data from northern New England. The Army Military Affiliate Radio System (MARS) and NWS have taken a first step in avoiding such problems in future. Army MARS has licensed a club station at the NWS forecast office at Albany, New York, AAT2CAA. This facility will make it possible for hams with MARS licenses to operate directly from the forecast office when links are needed beyond VHF range. As 1998 proved, it happens. All three MARS services--Army, Air Force, and Navy-Marine Corps--participate in the National Communications System's Shared Resources system linking federal agencies. Even if another NWS office lacks a MARS presence, Albany's traffic can now be relayed either by voice or digital mode through some other federal agency when needed.--*Bill Sexton, N1IN*

New address for VK4 incoming QSL bureau: There's a new address for the incoming VK4 QSL Bureau. It's GPO Box 199, Wavell Heights, 4013 Brisbane, Queensland, Australia.

Pacific Section to hold first-ever section convention: The Pacific Section will hold its first ARRL-sanctioned Pacific Section convention in Honolulu, Hawaii, October 13, 8 AM to 4:30 PM (gates open at 7:30 AM), at Pearl Harbor Community Park, next to the *Arizona* Memorial visitor entrance. Tickets are $2 each ($5 for family). All clubs will be given space for a display for the cost of paid admission tickets for all exhibitors. Tables will not be supplied. There will be a picnic area, vendors, flea market, displays, technical talks, amateur exams, and a lunch wagon. Several forums are scheduled, including an ARRL forum with Pacific Division Director Jim Maxwell, W6CF, and others. For more information, visit the Koolau Amateur Radio Club Web site.--*Bob Schneider, AH6J*

SETI@Home tallies 593,000 computing years: The Search for Extra-Terrestrial Intelligence@home project has completed its 593,000th computing year. SETI@home analyzes radio signals from space for signs of alien signals and uses idle PC time to do the processing. Users install the software on their home computers, where the program works in the background--exploiting the computer's processing power when its owner isn't. SETI@home is able to marshal the power of hundreds of thousands of PCs, processing more information than any supercomputer. The world's most powerful computer, IBM's ASCI White, runs at a speed of 12 teraflops, or 12 trillion floating point operations per second, and costs $110 million. In contrast, the SETI@home project runs at about 15 teraflops and has cost only $500,000. No conclusive evidence of extraterrestrial intelligence has been found, but project scientists say there are signals that may prove to be extraterrestrial. The answer, of course, is *out there*. For information on SETI or to obtain a copy of the software, visit the SETI@home site, http://setiathome.ssl.berkeley.edu/.

Southeastern VHF Society Conference set: The Southeastern VHF Society holds its fifth annual conference April 20-21 at the Holiday Inn Select-Brentwood, Nashville Tennessee. The program includes presentations by antenna specialist L.B. Cebik, W4RNL, VHF author and *QST* columnist Emil Pocock, W3EP, and EME enthusiast Bob McGraw, K4TAX. For more information, visit the SVHFS Web site, http://www.svhfs.org/.



This image of Hiram Percy Maxim is from a 16-mm movie frame taken at George Eastman's home at the introduction of *Kodacolor* movie film in 1927. It's one of the few images that depict Maxim laughing.

TV program on HPM scheduled to air: The life of Hiram Percy Maxim will be profiled in original films and interviews Tuesday, April 3, 7:30 PM, on Boston TV station WCVB-TV5 on the *Chronicle* news magazine. The program deals with recently discovered early films of New England. "Few people today realize that Hiram Percy Maxim was not only the founder of the ARRL but also the founder in 1924 of the Amateur Cinema League," says program producer Art Donahue, KA1GGG. "He chronicled his own life in hundreds of 16-mm movies from 1924 until his death in 1936." Donahue says the films were donated by his daughter, Percy Maxim Lee and his grandson, Hiram Percy Maxim II, to the Hartford, Connecticut, Public Library and Northeast Historic Film in Bucksport, Maine. The program uses this footage along with interviews with Percy Maxim Lee, Perry Williams, W1UED, the ARRL's former Washington, DC, representative, and David Weiss of Northeast Historic Film. Also shown is Maxim Memorial Station W1AW at ARRL Headquarters. The program also can be seen on many cable systems throughout New England and the Canadian Maritimes.

Visalia International DX Convention contest dinner set: The Third annual International DX Convention contest dinner will take place Friday, April 20, 8 PM, in the Oak and Maple rooms at the Holiday Inn-Visalia, 9000 W Airport Drive, Visalia, California. (Note: This is *not* the main banquet on Saturday evening, and the fee is not included in the convention registration cost.) Prime rib, $29; chicken princess, $23. No reserved seats. Tickets are available only from Champion Radio Products, 888-833-3104 (toll-free); Fax 530-758-9062; e-mail dinner@kkn.net; or write Champion Radio Products, 1816 Poplar Ln, Davis, CA 95616. Credit cards, checks and money orders accepted. Credit card users should supply card number, expiration date, and name as it appears on credit card. Deadline to order is Friday, April 13. No tickets will be sold at the door. Tickets will be mailed only to those supplying an SASE with their order; otherwise, pickup at the door. There will be a short program on contesting, guest speaker, and door prizes.--*Bernie McClenny, W3UR*

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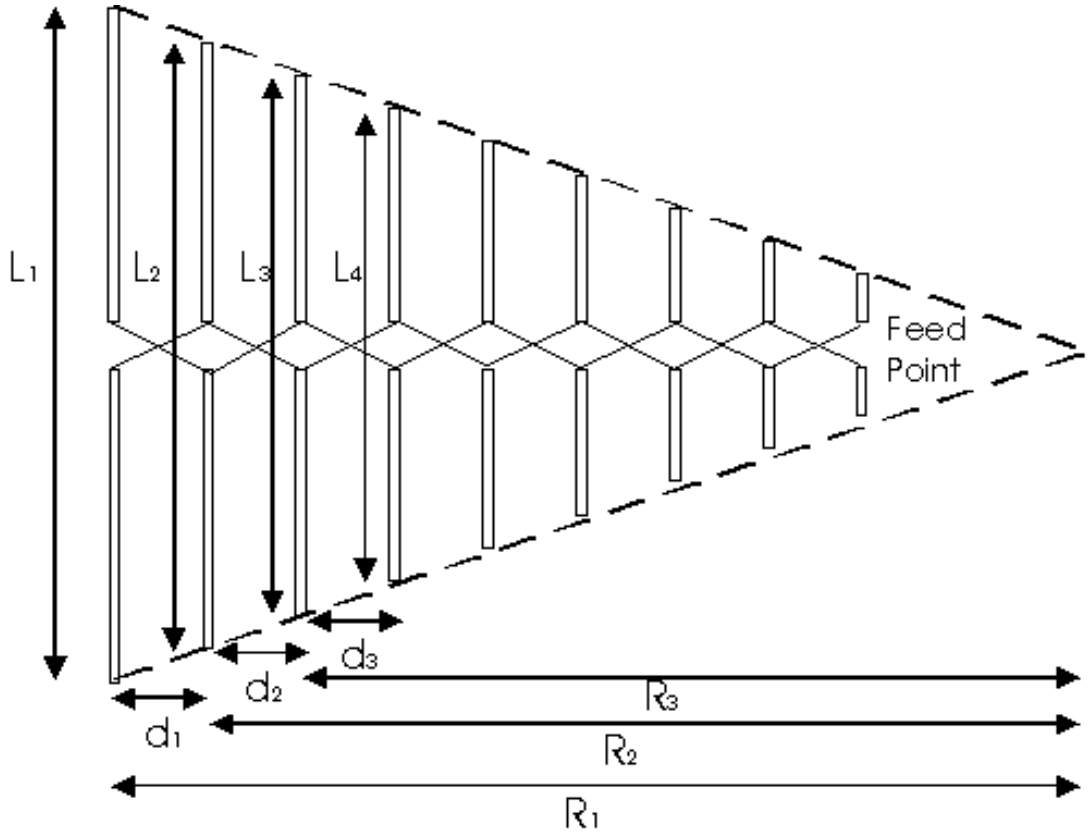
This is a Java applet which calculates the dimensions and spacings of the elements needed to build a log periodic antenna, given tao, sigma and the lower and upper cutoff frequencies.

All lengths and distances are given in meters. Look below for a diagram and a brief explanation describing what are all these numbers and parameters...



In a nutshell, The log periodic is a broadband antenna whose caracteristics vary as a periodic function of the logarithm of the frequency (with period Log(Tao)). As Tao approaches unity, we can neglect these response variations and regard the antenna as "frequency independant". Sigma is a spacing factor, and together with Tao, determine the directivity of the antenna.

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Long Wire Antenna

The long wire antenna is used as a field expeient antenna. To construct a long wire antenna you will need the following:

1. WD-1/TT, WD-1A/TT, or WF-16
2. Electrical Tape
3. 400-600 ohms terminating resistor
4. Insulators or material to construct field expedient insulators
5. Radio set
6. Knife
7. Suspension line
8. Measuring device
9. Compass
10. Signal Operating Instructions (SOI) with frequency and call signs
11. Large open area
12. Fabricated suspension (tree, Camouflage poles, etc.)
13. Calculator (optional)

To erect the antenna follow these steps:

1. Determine azimuth from the transmitting radio to the receiving radio.
2. Determine the length of the wire using the following formula:
    - **NOTE:For a quarter wave antenna divide 234 (contant) by the operating frequency in this case 9.00 MHz (234/9=26.0 ft.)**
    - **For a half wave antenna divide 468 (constant) by the operating frequency, in this case 18.00 MHz (468/18.00=26 ft.)**
    - **For a full wave length antenna divide 936 (constant) by the operating frequency , in this case 8.35 MHz (936/8.35=112 ft.)**
3. Measure and cut the antenna to the desired length
4. Split the WD-1 at the near end, attach one piece to the bottom end of a whip antenna and screw it into the antenna connector on the RT. Connect the piece to a screw on the RT case for a ground.
5. Suspend the antenna antenna using one of the following methods:
    - Run the antenna wire from the radio set to a nearbby tree. Attach the end of the antenna wire to an insulator. Attach a wire or rope from the other end of the insulator to the tree and tie it off.
    - Use a pole that is at least 9 feet high to suspend the antenna in the same manner. However you will have to use guy ropes to secure the pole.

The antenna is bi-directional. To make a uni-directional antenna place a 600 ohm resistor across the insulator.

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Last updated undefined

From: jpd@space.mit.edu (John Doty)
Newsgroups: rec.radio.shortwave
Subject: Re: SWL longwire
Date: 23 Dec 1996 19:49:01 GMT
Organization: MIT Center for Space Research

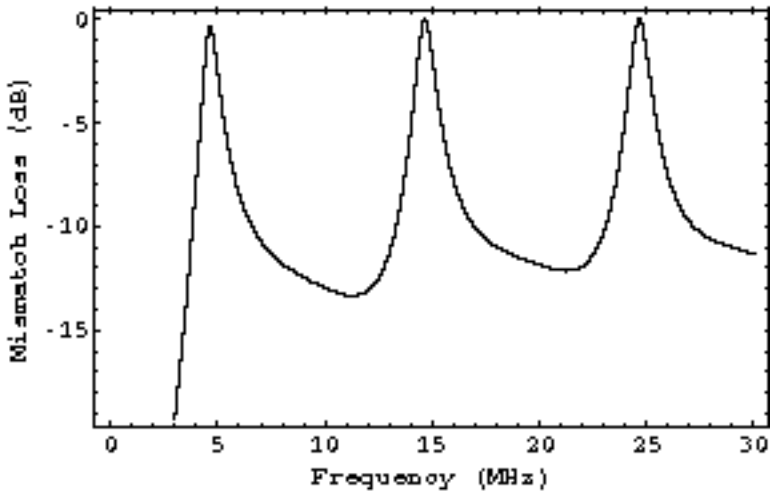
In article <9612182335114148@mogur.com> len.anderson@mogur.com (Len Anderson) writes:
> TV>> I wonder if a longwire balun would help match the impedance & provide a
> TV>> better signal?

> No, it will (primarily) change only the magnitude of the antenna
> impedance over frequency. Some bands will have more sensitivity than
> other bands. The antenna tuner will take care of that.

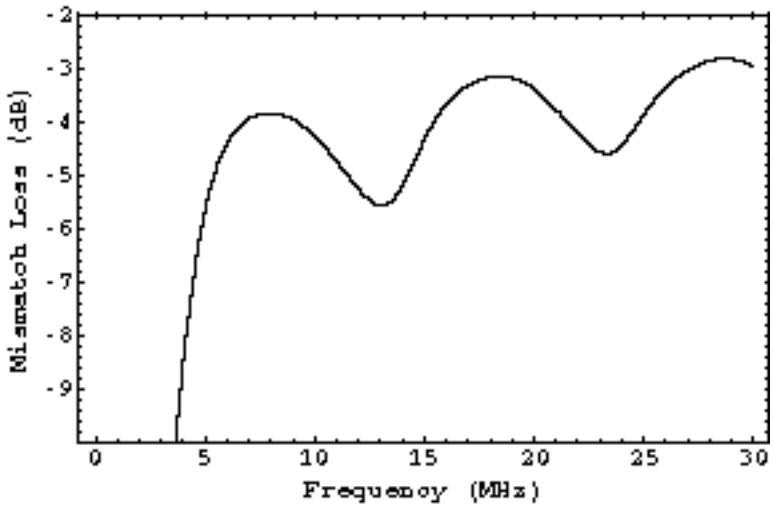
Actually, a fixed matching transformer can dramatically reduce the wild swings in antenna efficiency that a coax fed wire antenna exhibits. Let us calculate:

The following graphs are based on a 15 meter vertical antenna, fed at ground level, using a conical approximation. The antenna's characteristic impedance is assumed to be 620 ohms, which is typical for a thin wire. For more on the conical approximation, see Chapter 8 of "Antennas" by John D. Kraus (McGraw-Hill, 1950).

The first graph is for an antenna fed directly from 50 ohm coax. The horizontal axis is the frequency in MHz, the vertical axis is the mismatch loss in dB. The well known "quarter wave" resonances near 5, 15 and 25 MHz are visible as sharp peaks where the mismatch loss closely approaches zero.



The second graph assumes a matching transformer with a 9:1 impedance ratio at the feedpoint, presenting the antenna with a load resistance of 450 ohms. At most frequencies, the mismatch losses are considerably lower for this case. The variation in the mismatch loss is also reduced:

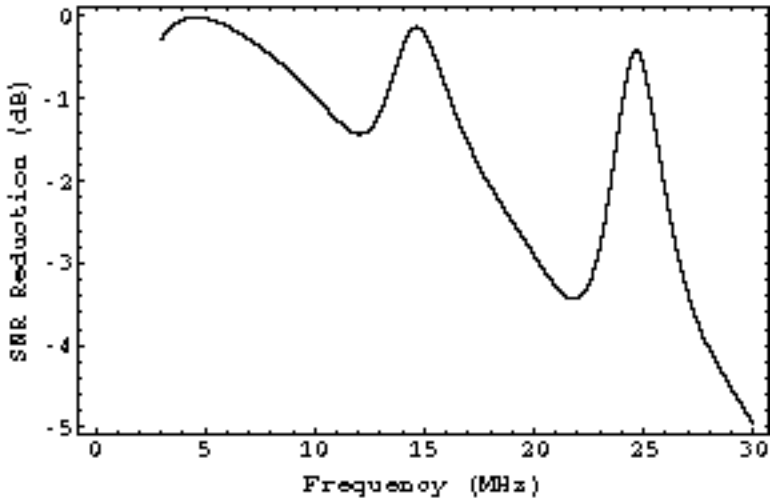


Well, so what? In the absence of interference, the signal to noise ratio is the main determining factor for the audio quality of the signal. The mismatch loss affects both signals and noise, so if the receiver was

noiseless the losses would not affect the signal to noise ratio. Real receivers, however, are not noiseless: if the loss is too high, receiver noise will become dominant, and overally system sensitivity will suffer.

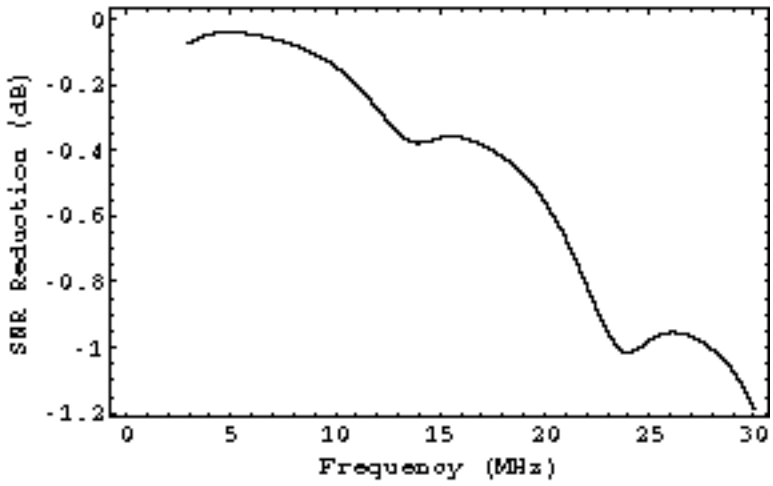
The following results assume cosmic noise of 29 dB above thermal at 10 MHz, declining with increasing frequency at -23 dB per decade. No man made or atmosperic noise is assumed. I assume a receiver noise figure of 10 dB.

First, here is the signal to noise impact of the mismatch losses for a 50 ohm coax feed without a transformer:



Losses in signal to noise of 3-5 dB are likely to be noticeable. The largest impact is in the quiet bands above 15 MHz.

On the other hand: the loss in signal to noise with a 450 ohm feed is much smaller:



You are unlikely to be able to notice losses in signal to noise in this range.

The results depend on the assumptions. A real longwire isn't usually vertical: this tends to degrade its performance a bit at the low frequency end, while improving it at high frequencies. This is good, because in the model the signal to noise is declining as the frequency increases: the increase in performance cancels part of this.

No man made or atmospheric noise is included. If they are significant, the precision of the match becomes less critical. Man made noise can be significant at any frequency, but atmospheric noise is more significant at the lower frequencies.

A receiver noise figure of 10 dB is mediocre for a solid state receiver or a tube receiver with a triode RF amplifier. Tube receivers with pentode RF stages may be a bit worse than this, and something like a Hallicrafters "Sky Buddy" (no RF stage, pentagrid converter) might have a noise figure >30 dB. The better (smaller) the noise figure, the less you have to worry about matching. Sky Buddy owners will want to tune their antennas very carefully.

I haven't included cable losses here. These are not terribly important unless you're using an ATU at the receiver end. If you are, using a fixed transformer to get the match roughly right at the antenna end will reduce the cable losses, because cable losses increase with increasing SWR.

My own experience concurs with the results of this theoretical analysis (or I wouldn't be writing about it: I'd be trying to figure out what was wrong!). I have experienced "deaf bands" with coax fed antennas lacking matching transformers, but my transformer-fed antennas work well across the HF spectrum (and

even down to longwave). I don't bother with an ATU.

See "Low Noise Antennas" at "http://www.anarc.org/naswa/badx/" for more on matching transformers and on keeping conducted noise from contaminating a "longwire antenna".

John Doty "You can't confuse me, that's my job."
jpd@space.mit.edu

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Last modified: *March 10, 1998*
E-mail your comments and updates to Steve Byan <smb@world.std.com>

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LOW NOISE RECEIVE-ONLY COAX-LOOP ANTENNAS

for 160 - 10 meters HF bands

Station WN6F listening post

Last Update: 29 March 2004

After doing battle with locally-generated EMI / RFI for nearly two decades, I decided to do something about it. This antenna's main advantage is that it packs a lot of low-angle directivity into a small package. This makes nulling out many urban noise sources easy while enjoying good medium to high-angle skywave reception.

NEW !! (If you are interested in an easier-to-construct wire-only loop that has better performance, check out my general-purpose wire-loop page).

The directional small loop is also useful for significantly reducing static-crashes on the lower HF bands at times; this depends on mother nature of course. The larger loop is for 80 meters and the smaller loop is for 40 meters.

The project is described in four parts:

- General construction and usage (this page)
- Operational analysis of this small loop
- Alternate loop winding methods
- Miscellaneous notes and experiments

They are easy to build, and the only tools you'll need are wire cutters, some coax, and a tuner. You can make them for yourself or friends in a matter of minutes. Part of my design goal was to use a minimum of easily available parts. You won't need a well-stocked junkbox or any exceptional mechanical skills to build this antenna.

Winding the loops out of coax (providing a main antenna loop and an interior coupling loop) in the following fashion is very convenient and has six purposes:

1. The small wavelength outer-conductor loop (at 0.10 or smaller wavelength in circumference) provides the horizontal low-angle bidirectional directivity with deep nulls.

2. The gap in the braid at the top center of the loop allows common-mode reception of signals (both wanted and unwanted) on the outer skin of the loop braid to transition to the inner-skin of the braid. Rotational directivity nulls out the unwanted noise.

3. The current on the inner-skin of the braid now acts like a normal transmission line via the differential-mode to the center conductor.

4. This specially-wound loop acts as it's own balun, which is necessary when using unbalanced coax as the feedline. This helps to ensure that the feedline does not become part of the antenna. If it did, you'd lose much of your directionality, or have very strange nulls.

5. The large amount of distributed capacitance makes the loop somewhat

resistant to unbalance and detuning from hand-body-object capacitance than more traditional balanced wire loops.

6. The larger outer conductor diameter of the braid increases the sensitivity of the loop over that of a small-gauge wire.

Prove it!: Attach a clamp-on ferrite RF cable choke on the feedline and arms of the loop. When chokes are placed anywhere on the arms of the split-braid loop, severe detuning and attenuation takes place. When placed anywhere on the feedline after the loop output, no detuning or attenuation occurs.

HF loops built in this way are typically used for direct wave direction-finding. I'm more interested in using them as near-field noise-nullers to enhance the signal-to-noise ratio of skywave signals. This helps make up for their inherently lower gain than more traditional long wire antennas. A balun-effect is created by the way we wind the antenna back onto the feedline to help prevent common-mode noise ingress and maintain an accurate directional pattern.

2004 update: See below for smaller TWO-TURN Loops!

I'm using these antennas indoors right next to the operating position to make it easy to rotate and tune. Remote mounting is certainly possible, preferably with an antenna tuner or preamp located near the loop, although in that case I'd advise using a plain wire loop if you want to keep the tuner near the operating position.

As an example, let's build a single-turn 0.10 wavelength loop for 40 meters: (it will also function well as a .05 wavelength loop on 80m with retuning)

WHAT YOU'LL NEED:

- For 40 meters, you'll need about 15 feet of coax (loop and some extra length for tuner connection, plus pigtail wrap.) I prefer to use 50-ohm RG-58. For convenience I picked up a 50-foot jumper at Radio Shack (#278-971A) which is already connectorized so I could build a few of them. You could also use 75-

ohm coax without any difficulty. Try to get the lowest capacitance cable that you can find.

- An antenna tuner can be used to tune the loop system to resonance. I've used L-type tuners and T-type tuners without any problems. I'm currently using an MFJ-16010 random-wire tuners which are small and use two SO-239 connectors which makes them easy to hook up and mount to a mast if desired. Note that on 160 meters, you'll need a lot of inductance, and the 16010 model doesn't seem to have enough when used with coax-loops.

- Optional: A 10 dB gain or more preamp can be used to help bring up the signal level. If your receiver has a preamp, and all it seems to do is bring up the noise level along with the signals, you are in for a treat. With this directional antenna, the preamp finally becomes useful in amplifying the signals, and not the noise.

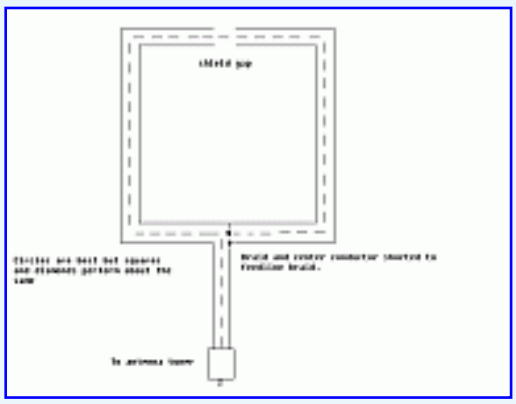
Use the largest coax diameter you find practical. I use RG-58 since it is easier to work with, but the larger conductor size of RG-8 / 11 should perform a little bit better.

LOOP SCHEMATICS:

My best artistic attempts at drawing the schematics.

Click on the images for a larger picture.

(Internet-Explorer users: IE has an option that automatically reduces the size of graphics. If the enlarged schematics look distorted, be sure to view them full-size, typically with F11 Key)



HOW TO MAKE IT:

(Click on any of the images to enlarge them.)

1) First, at one end of the coax, let's short the braid and center conductor together with a pigtail. RG-8 was used to make it easier to see, although I like to use the smaller RG-58 or RG-6. After removing the outer covering, I bunched the braid back a bit to allow me to cut into the inner dielectric and then just pushed it back down when I was done. Take care not to cut or score the inner conductor with the knife, especially when using the smaller cables; when you wrap the pigtail around the braid in a later step, it has a tendency to snap at the cut or score when you wrap it.

(It might be easier to build the loop by following these instructions backwards if you already have a connectorized piece of cable.)

2) We need to create a 1-inch gap in the coax shield. Measure 7 feet from the shorted end, and make a small cut mark on the coax. A half inch on either side of the mark, cut through the outer covering exposing the braid. Then cut the braid out to expose just the inner dielectric. Try not to cut into the inner dielectric as this will mechanically weaken the antenna when you eventually mount it. The photos show some exposed braid at the gap just for clarity, but my antennas just have a clean cut. Note that this is just a true gap, you do not short anything at the gap! (Make sure there aren't any fine strands of shield wire still connected across the gap after you cut it!)

KEY POINT: The gap allows common-mode reception of radiated energy (both wanted and unwanted) on the outer braid skin to transition to differential-mode currents on the inner braid skin, which is coupled to the center conductor via differential-mode. Rotation nulls the unwanted common-mode energy exposed to the arms of the loop.

3) Measure another 7 feet from the center of the gap, and remove about 1 inch of the outer coax covering ONLY. You should just have some exposed braid.

4) Take the shorted pigtail end of the coax, and wrap it around the exposed braid from the previous step. This is the weakest point of the loop mechanically. Try not to make a radical bend at the point where the brittle center conductor is immediately shorted to the braid; it has a tendency to break if you do. Wrap at a point just a little bit further down the pigtail to avoid breaking the center conductor.

KEY POINT: This finishes the self-forming balun and helps ensure that the unbalanced feedline does not become part of the balanced loop, thereby unbalancing it, and turning the loop into a noisy random wire.

(NOTE: don't be tempted to use an ordinary "tee" connector here. It won't work properly.)

You now have a highly directional low-band 40 meter loop and the rest is a small length of feedline that goes to the input of a tuner. Of course the jumper from the tuner output to the receiver can be any convenient length.

For 160 through 40 meter operation, you can mount the tuner anywhere from 6-inches to 6-feet away from the output of the loop. This allows you some flexibility if you want to keep the tuner on the desk and not on the mast. For smaller single-turn loops on 20 meters and above, try to keep the tuner no more than 6-inches away from the loop output - the closer to the loop output the better.

Note: Here is a very small loop for overall visual reference.

To construct a 20 meter loop, try a loop circumference of 7 feet. The gap is not critical for any band, and can stay at 1 inch. For 80 meters, try a loop circumference length of 27 feet or so. For 20 - 15 meters, try a 3-foot circumference; I'm amazed that this actually works as well as it does.

Note that when we cut loops for 0.10 wavelength, they will also perform well at half their frequency as a 0.05 wavelength, although sensitivity will be down somewhat. The positive tradeoff is that the nulls will be VERY deep at 0.05 wavelength.

Design Your Own!

UPDATE: MAJOR ERROR IN MY EARLIER DESIGNS!

My apologies go to earlier readers of these pages. I incorrectly assumed that since we are using coax as antenna elements, that they would be subject to a cable's velocity factor. Since the loop element is really the outer braid skin, **it is not subject to velocity factor.** Only the inner differential-mode of coax is subject to velocity factor, and NOT the common-mode outer surface conductor of the braid, which is what we are using as the antenna!

My standard formula for calculating the loop circumference for the frequency of interest at 1/10th wavelength *without any VF*, is:

(1005 / freq mHz) * 0.10

TUNING NOTES

This formula should get you in the ballpark, although if you can't get a definite peak with your tuner on the lowest frequency of interest, it is likely that the antenna is too long, or the distributed capacitance of the cable used is too high for your tuner to handle. Try using a more advanced tuner, or reduce the loop circumference length.

You can cut a new loop with a cable that has lower capacitance specs, or just make the existing one shorter. Fortunately it is easy to make a pre-cut loop shorter. Let's cut off two feet from the existing loop circumference as an example:

- Unwrap the shorted pigtail from around the exposed braid. Cut away a foot from this arm of the loop and create another shorted pigtail.
- Measure a foot from the center of the old exposed braid, and cut a new 1-inch exposure of braid.

- Wrap the new pigtail around the new exposed braid section and re-shape the antenna into the smaller shape. Tape up or heat-shrink the older exposed braid section.

- Try to be as accurate as possible and strive for balance when re-sizing the loop. Balance is important!

USAGE:

You can create the loop with any shape you want, but circles are best, and squares and diamonds do nearly as well. Triangles, and extreme rectangles don't do as well, but don't let that stop you. The key is to enclose as much area as possible. For mounting, any sort of cross-brace should work, but I've also placed them on the back of non-metallic doors, hung them from ceilings, etc.

Tune the loop-system (loop and feedline) to resonance with the tuner. Since loops are high-q, you should hear a definite peak in the background noise. Better yet, tune to some locally-generated noise and peak on that. As you change each setting of the inductor, rotate the tuner capacitor(s) s-l-o-w-l-y.

The directivity pattern is a bidirectional figure-8 pattern at low angles, with maximum gain in the plane of the loop, and the nulls are broadside to it. It is just the **opposite** of an ordinary dipole. The loop seems progressively more omnidirectional to medium and high skywave angles.

If you want to see the directional azimuth and elevation angles, check out my general-purpose wire-loop page. The patterns are almost exactly the same as compared to these coax-built loops.

You don't have to have the loop perfectly vertical, use whatever angle you need to null the local noise or storm-related static crashes - skywave signals will still be heard well. In one case, I have a noisemaker downstairs, so I had to rotate and tilt the loop at about 45 degrees to null it - it didn't attenuate the desired band signals at all. At times I've been able to track a storm all night and reduce the static crashes, and then sometimes not depending upon how localized the storm is.

You might notice that even though you've nulled out the noise, signal levels on a tuned loop seem to be about 2-to-3 S-units down from a standard 1/2-wave dipole mounted up at least a quarter-wave above ground. This is normal for a *tuned* small

loop given its short wavelength. You can compensate with a preamp, but you might find that with the noise level so low, you won't need one. If your normal antenna isn't mounted up at least a quarter-wavelength, the comparative loss might be even less!

Note: many modern radios that have built-in selectable preamps might have much lower sensitivity with the preamps off than other radios that don't have any preamps at all. This is a good design feature. Accordingly, with radios that have selectable preamps, it is likely you'll want to run with at least the first preamp turned ON.

The important point to remember is that your noise floor will be so much lower than your old noisy antenna that it more than makes up for brute signal strength.

Connector Quality Issues

Make sure your connections are snug. If you lose the ground, you'll end up with a random-wire because you've lost the electrical balance. I mention this because as you insert or remove the connector from the tuner or receiver, when only the center-conductor is making a connection, you'll hear noise or signal level come up a bit. Unfortunately, you will have lost all the directivity that this antenna provides. Keep those connections tight.

A caution about crimped PL-259 connectors: sometimes the crimp is so tight that when you screw them fully into the chassis jacks, they don't actually make good contact with the chassis-connector center conductor. A **VERY GENTLE** squeeze with a pair of pliers on the center-conductor pin will help ensure good physical contact. I've run into this problem a lot with cheaper cables and sometimes it is not immediately apparent because you can still receive signals (poorly) due to capacitive coupling of the center pins.

Another drawback of cheaper connectors is that when you mate the dielectrics of the cable plug and chassis jack together, they sometimes don't allow the shells to make contact, especially if the "teeth" are short. You might only have a very poor shell-to-chassis connection via the rubbing of the cable's screw ring to the shell.

TWO-TURN LOOPS!

To turn a large loop into something more manageable size-wise, try a continuous TWO-TURN loop. Each turn should have a 1-inch gap at the top of the turn, and at the bottom of the turn you attach the exposed braids and shorted pigtail together.

Turn spacing: you might want to put a coax-diameter's width of space between the two turns. I did this by winding one turn on the front of the mast and the other turn on the back.

(Click on any image to enlarge)

For example, my 40 meter single-turn diamond loop has a 14-foot circumference with an approximate 4-foot diameter.

I turned it into a two-turn loop by starting with a new 14-foot length of coax, and depth-winding a second continuous turn around a 2-foot circular diameter instead of a square 4-foot diameter. At the top of the loop are my two turns with the 1-inch gaps at the center. At the bottom output of the loop I made sure the exposed braids and shorted pigtail are connected together. From this point, I left about 6 inches of coax terminated with a PL-259 connector which attaches to my portable MFJ-16010 tuner.

(Note: I recently placed the tuner on the desk with a 3 foot jumper instead since this loop was designed for the lower bands. For smaller single-turn loops designed for 20 meters or higher, the length of feed from the output of the loop to the tuner should not exceed 6 inches - the closer to the output the better.)

Try to make all your measurements, cuts, and braid connections as symetrical as possible. Making the loop balanced is key to good noise rejection.

Here is how I made it:

- I used a 15 foot length of coax to allow for a short 6-inch run to the tuner and

some left over to make the shorted pigtail.

- Six inches from the end of the connector, make a 1-inch cut around the coax to expose only the braid.
- Measure 3.5 feet from the center of the exposed braid, and cut a 1-inch gap (1/2 inch on both sides of the mark) through the braid to expose the dielectric.
- Measure 3.5 feet from the center of the exposed dielectric, and cut around the coax for 1 inch to expose only the braid. (1/2 inch on both sides of the mark.)
- Measure 3.5 feet from the center of the exposed braid, and cut a 1-inch gap (1/2 inch on both sides of the mark) through the braid to expose the dielectric.
- Measure 3.5 feet from the center of the exposed braid, and short the shield to the center conductor. Leave about 4 extra inches or so of this shorted pigtail.
- Wrap the coax into a two-turn loop with the exposed braids parallel to each other at the bottom, and the exposed dielectric gaps at the top. Use tie-wraps, tape, etc.
- Wrap the shorted pigtail around the centers of the exposed braid sections at the bottom so that all the braids are in contact. Solder or tape together.
- Put the loop on your support, attach your tuner to the connectorized end and enjoy!

The results of this two-turn loop are great! Signal output levels between my single and reduced-area double-turn loops seems about the same.

Casual MF BCB DX'ing and LF

If you bypass the tuner, you might be able to use a loop cut for 80 or 160 meters in an untuned mode for casual MF BCB and LF reception. It is handy to null out the local broadcasters.

With the loop untuned, I've finally heard low-frequency CW beacons. So there IS life down there! For serious use, I'd definitely recommend specialized antennas for these bands, but at least an untuned loop might whet your appetite to go "lowdown".

Other Sources of Info

I am indebted to all the Amateur Radio operators, SWL's, BCB DX'ers, VLF'ers and other experimenters that provide me with information and encouragement with this project. They participate in the **Yahoo!® Loop group** and also on the **rec.radio.amateur.antenna** newsgroup. All errors and misinterpretations belong to me however ...

If you're interested in loops, whether they be shielded, unshielded, magnetic, loopsticks, etc, you can find a great discussion forum for loop antennas on Yahoo!® by clicking here.

Be sure to visit the K7ZB site with his pics and nice build of this loop, along with other interesting antenna solutions.

- General construction and usage (this page)
- Operational analysis of this small loop
- Alternate loop winding methods
- Miscellaneous notes and experiments
- Easy-To-Build Small Wire Loops!

since 1 June 2003

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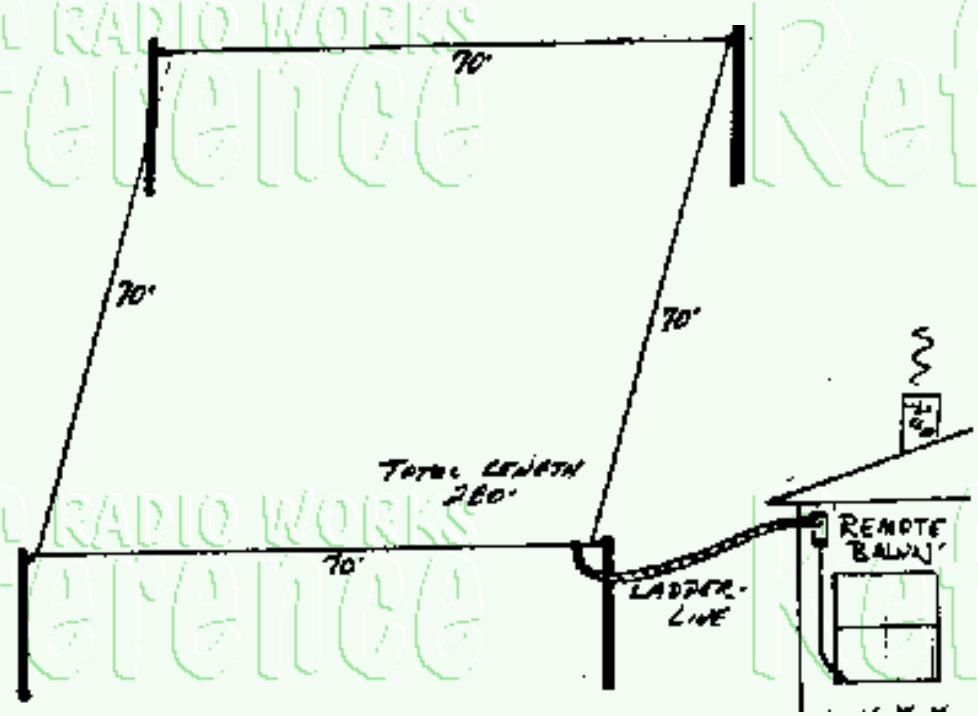
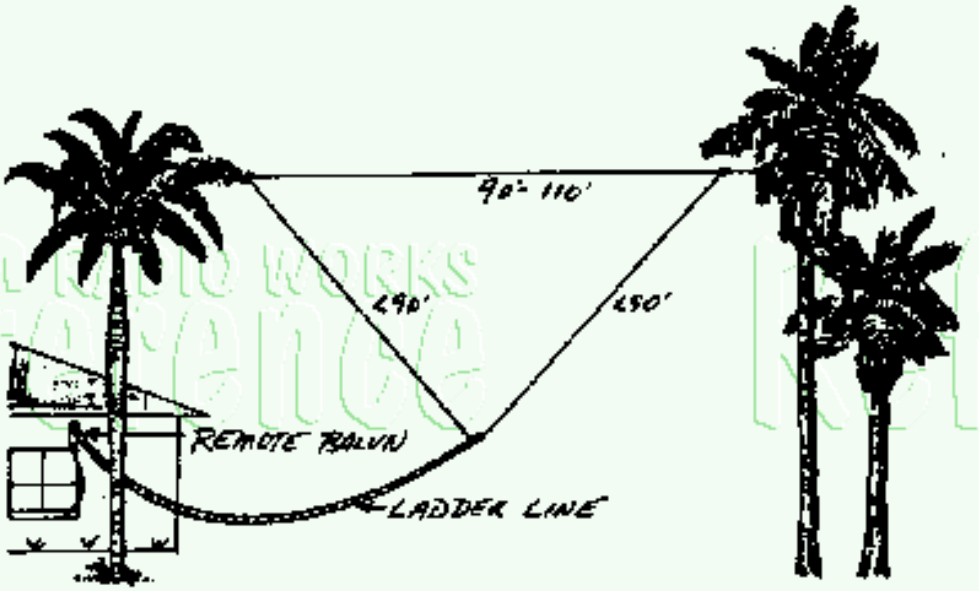
LOOP ANTENNAS





Full-wave loops are very popular antennas. They are especially useful on 80 and 40 meters where they perform well at modest heights. These are closed loops that are one full wavelength long. Horizontal loops may be fed at any convenient spot. For best performance, make your horizontal loop into a square, especially if it is to be used on several bands.

The Vertical Loop is a good DX antenna. The shape can be a circle, square, rectangle or a triangle. The larger the area of the loop the better it will work. Feed square and rectangular loops at a corner. For best results, triangular loops should be supported apex-down. This puts less of the antenna parallel with the ground and increased the effective height. Feed triangular loops either at a corner or in the case of apex-down loop, at the apex.

Use ladder line and a wideband transmatch (a naturally balanced tuner, like a Johnson Matchbox) for multiband operation. The RemoteBalun 4 is recommended if you will have problem getting ladder line to the operation position. Multiband operation is possible when feeding the loop with coax. The losses will be slightly higher, but the convenience of the coax may be worth the slight signal loss.

The design frequency, the feedpoint impedance, will be between 80 and 150 ohms. Coax fed loops will usually have an SWR between 2:1 and 3:1. You may feed this antenna with a 4:1 balun. If the loop is in the shape of square or large rectangle, the SWR can be below 2:1, but will not get much below 1.5:1.

If you decide to feed your loop with coax, I'd suggest using RG-8X or RG-213 and a high power, high performance 1:1 or 4:1 Current-type balun. Experiment with full-wave loops. You may find them to be excellent multiband antennas.

Antenna wire can be #14 hard-drawn antenna wire. Use #12 wire for large loops on 160 or 80 meters.

Center Frequency	Total Length in feet	Center Frequency	Total Length in feet	Center Frequency	Total Length in feet
3.5 MHz	287	10.12 MHz	99	28.0 MHz	35' 10"
3.6	279	14.0 MHz	72	28.5	35' 3"
3.7	272	14.2	70	29.0	34' 8"
3.8	264	18.12 MHz	55' 6"	29.5	34' 1"
3.9	258	21.0 MHz	48' 10"	---	
7.0 MHz	143	21.2	47' 6"	---	
7.2	139	24.93 MHz	40' 4"	---	

With this particular antenna, a RemoteBalun 4 and Ladder Line are used to permit multiband operation. If you have a naturally balanced tuner (i.e. a Johnson Matchbox), the RemoteBalun 4 is not needed. Click on the Balun Index do check out the the RemoteBalun's application notes.

Balun Index TOP Jim's Notebook Index

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Description

Category: Shortwave

Discussion of loop antennas on all bands including ferrite loops, air core, all types. All bands SHF, VLF, LF, MF, HF, VHF, UHF, etc. Shortwave, longwave, receivers, transmitters, all fair game here.

Trying to resurrect the old LoopAntennas list without the melodrama. Please post links, files, etc. Please write a good description of files and links when you post them. I saved all the links, we'll have to build up the archives again.

All new members are moderated until you post once or twice, or if I recognize you. Spammers will be handled with minimum fuss.

What's New

Nothing new within the last 7 days

Message History

	Jan	Feb	Mar	Apr	May	Jun	Jul	Aug	Sep	Oct	Nov	Dec
2004						131	21	49				

Group Email Addresses

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Founded: **Jun 7, 2004**
Language: **English**

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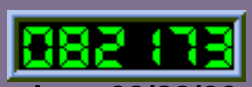

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g4fon

Welcome to my web site, my name is Ray Goff.

I was brought up in Farnborough, Hampshire and first became interested in radio at the tender age of eight while staying with my older cousin for a weekend and we built a crystal set together. This interest developed throughout my teenage years, to the extent that in my late teens I was running a lucrative business mending all the neighbourhood TV's and Radios.

At the same time I became interested in short wave listening after my father produced an old German Torn Eb TRF receiver, which he soon augmented with an R1155 set. This was the end of the 1960's and I had already joined the local Farnborough and District Amateur Radio Society.

In 1969 I started evening classes to pass my RAE exam. The course was held in a local school under the tutelage of John Hardy, G3KND. I passed the exam at the first attempt in May 1970, but was determined to hold out the for the 12 WPM Morse

test and a full license.

By 1973, while at university, it became clear that despite my best efforts, including help from another student, G3VMO, I was not going to pass the test in the foreseeable future, so I applied for a Class B, VHF only privileges license and received the callsign G8HMH.

Briefly back in Farnborough again, I joined another evening class led by Colin, G3XUU, to learn Morse code and in November 1976, finally passed the Morse test at a coastal radio station in Southampton and upgraded to my Class A license, G4FON.

From 1981 to 1984 I was G4FON/W9 while living in Chicago and then G4FON/W1 until 1989 while living in the Boston area.

In 1989 I returned to the UK and now live in the centre of Oxford which presents problems for HF operation. After a number of years of inactivity, I can now be found on 20, 30 and 40 metres CW, usually in the QRP section of the band.

This website reflects my main interests; QRP CW operating,

homebrew construction and the contesting. You are encouraged to look around and make use of anything you find. I would love to hear about anything you find useful, so drop me an email

On this website you will find a CW trainer which I wrote to build my CW speed back up, construction notes for homebrew equipment and aerials and a logging programme for the Palm OS. You will also find my latest programme, a Windows based CW Contest trainer.

More recently, I visited the Farnborough and District Amateur Radio Society for a meeting and renewed some old friendships and I am an active member of both the Oxford and District Amateur Radio Society and the Newbury and District Amateur Radio Society.

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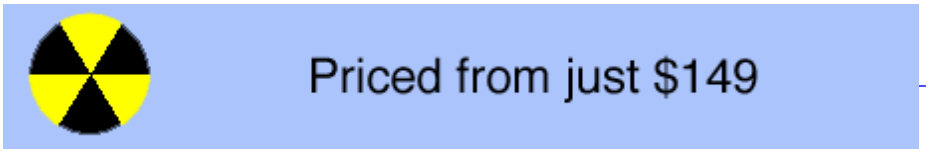

RF Toolbox

RF Toolbox is an antenna design and electronics/electrical tool package. It takes you step by step through the design of the following types of antennas:

- Dipole
- Fat Dipole
- Yagi
- J-Pole
- Super J-Pole
- Log Periodic
- Cubic Quad
- Vertical

It also performs the following calculations:

- Coil Design
- LC Filter Design
- Transmission Line Loss
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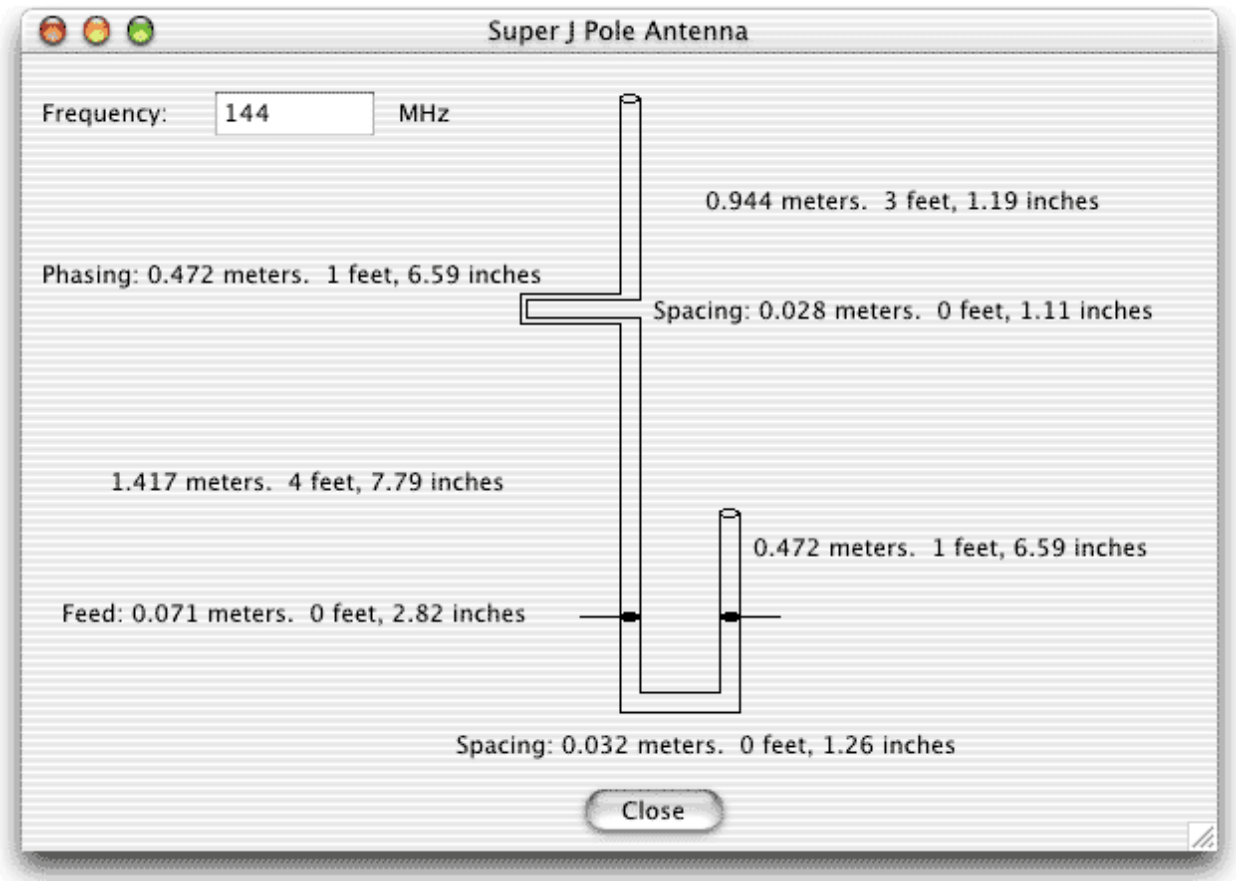
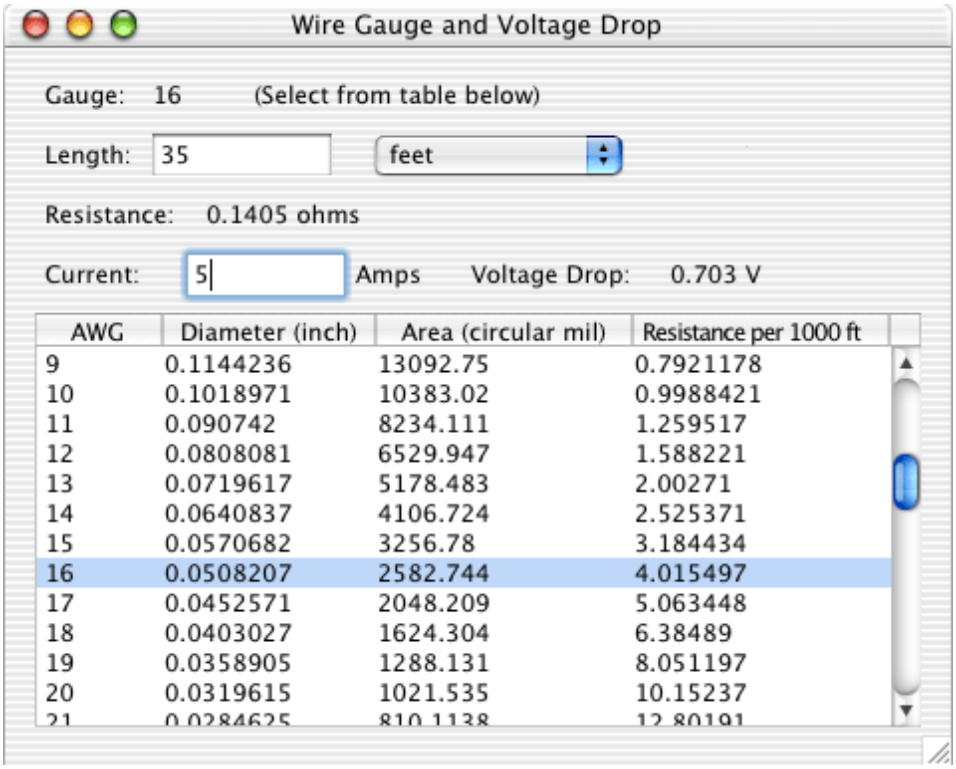
- Pi Matching Networks
- Impedance Calculations
- Wire Gauge and Resistance and Voltage Drop
- Wire Inductance Calculations





AWG	Diameter (inch)	Area (circular mil)	Resistance per 1000 ft
9	0.1144236	13092.75	0.7921178
10	0.1018971	10383.02	0.9988421
11	0.090742	8234.111	1.259517
12	0.0808081	6529.947	1.588221
13	0.0719617	5178.483	2.00271
14	0.0640837	4106.724	2.525371
15	0.0570682	3256.78	3.184434
16	0.0508207	2582.744	4.015497
17	0.0452571	2048.209	5.063448
18	0.0403027	1624.304	6.38489
19	0.0358905	1288.131	8.051197
20	0.0319615	1021.535	10.15237
21	0.0284625	810.1138	12.80191

The cost of buying RF Toolbox is only $14.99!
Buy your copy online, and get your registration code in minutes!

RF Toolbox is available for both Macintosh and Windows systems!

Email your comments and questions to info@blackcatsystems.com

Last modified October 18, 2003

Welcome to W2BRI's Magnetic Loop Site

Magnetic Loop Antennas for HF & Above

Last Updated October, 18th 2003
New Review of the AOR ARD9800 Digital Voice Modem!

New!

W2BRI gets the digital voice bug. Aside from my love of magnetic loops, I have been focusing my attention towards a new mode of communication for ham radio, digital voice.

Here is a link to my Digital Voice site.

My **new review** of the AOR ARD9800 Digital Voice Modem is ready for your reading pleasure!

Introduction

My current passion in radio has become the Magnetic Loop Antenna, also referred to as STLs, small transmitting loops. This site is dedicated to these antennas, and I strive to make this

Site Menu

A. Magnetic Loop Antenna FAQ

In this section you will find theoretical and practical information regarding mainstream beliefs, practical observations, and down to earth theory about these antennas. This FAQ will answer many major questions.

B. My 80 Meter Magnetic Loop Antenna

Find out about my largest magnetic loop for 40 and 80 meters. This page tells the story about how I came to learn about magnetic loops, and my experiences with this wonderful antenna.

---

the most comprehesive site possible. Magnetic Loops should not be confused with full size resonant loops commonly found in many radio operators backyards. The Magnetic Loop antenna has some special properties which distinguish it from many antenna designs like the dipole.

A major advantage of the magnetic loop is its high efficiency and small size. The magnetic loop is typically smaller in circumference than 1/4 wavelength of the desired operating frequency. There are many more advantages I plan on describing on this site and some disadvantages. The antenna is ideal for restricted operating areas and for portable operation. They can work close to the performance of a full size dipole antenna (sometimes exceeding a dipole) and do not require radials. They can be mounted low to the ground, and will exhibit a high level of performance. However, they tend to be very expensive devices when bought commercially. They can be homemade with relative ease, and this site will teach you how to build your own if you so choose.

Loop antennas are elegant antenna systems, but they are far from simple. The Magnetic Loop is a complex antenna and will not simply work as a plug and play solution. If one desires using the loop antenna effectively, a decent understanding of the antenna's characteristics are crucial. A Magnetic Loop can work very well for an operator if they understand the different variables necessary to make it work well. My intention is to elucidate many of these issues on this site and make it easier for operators to use these antennas successfully. I also intend to discuss some of the myths related to loop construction, and hope to clear up some possible misconceptions others have encountered in with this interesting antenna.

My hope is to open up the technology to more operators. I will welcome phone calls and questions at 818-505-0222 or by email at brian@standpipe.com, but perhaps you'll find the

C. How to Build Your Own Magnetic Loop Antenna

If you wish to build a Magnetic Loop antenna and have questions about how to make it work, this section is for you. This five step project plan explains all the steps to creating your own magnetic loop from basic plumbing materials. The article features step by step instructions and how-to pictures to guide even the most basic beginner in Magnetic Loop construction. The article also features a section on capacitors and loop tuning. It is a must read.

D. Images and Notes from other Loop Builders

You are not alone. There are many others out there building and experimenting with Magnetic Loops. This section features great articles and pictures of other loop builders and their experiences. If you are serious about what's going on out there in the world of Magnetic loop, then this section is essential. If you are interested in submitting an article, please send me an email at brian@standpipe.com

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E. All About Variable Capacitors

Many of you have asked for this section and here it is! Where to buy your vacuum variables, what to watch out for, and how to use them. This section also covers air variables, sources to buy them, and additional info. Enjoy.

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answer to what your looking for on this site and build it yourself!

I am always looking for original artciles, images, and info on Magnetic Loops. Please email me at:

brian@standpipe.com

if you would like to submit content for the site.

If you are interested in the topic of project management, I write a monthly article which can be found at www.allaboutpojects.com

F. Loop Software Page

In this section you'll find some very easy to use loop modeling software. It's as easy as pie to plug in your desired parameters and see how well your antenna will perform theoretically. There is also a very cool homebrew capacitor design program. Check it out!

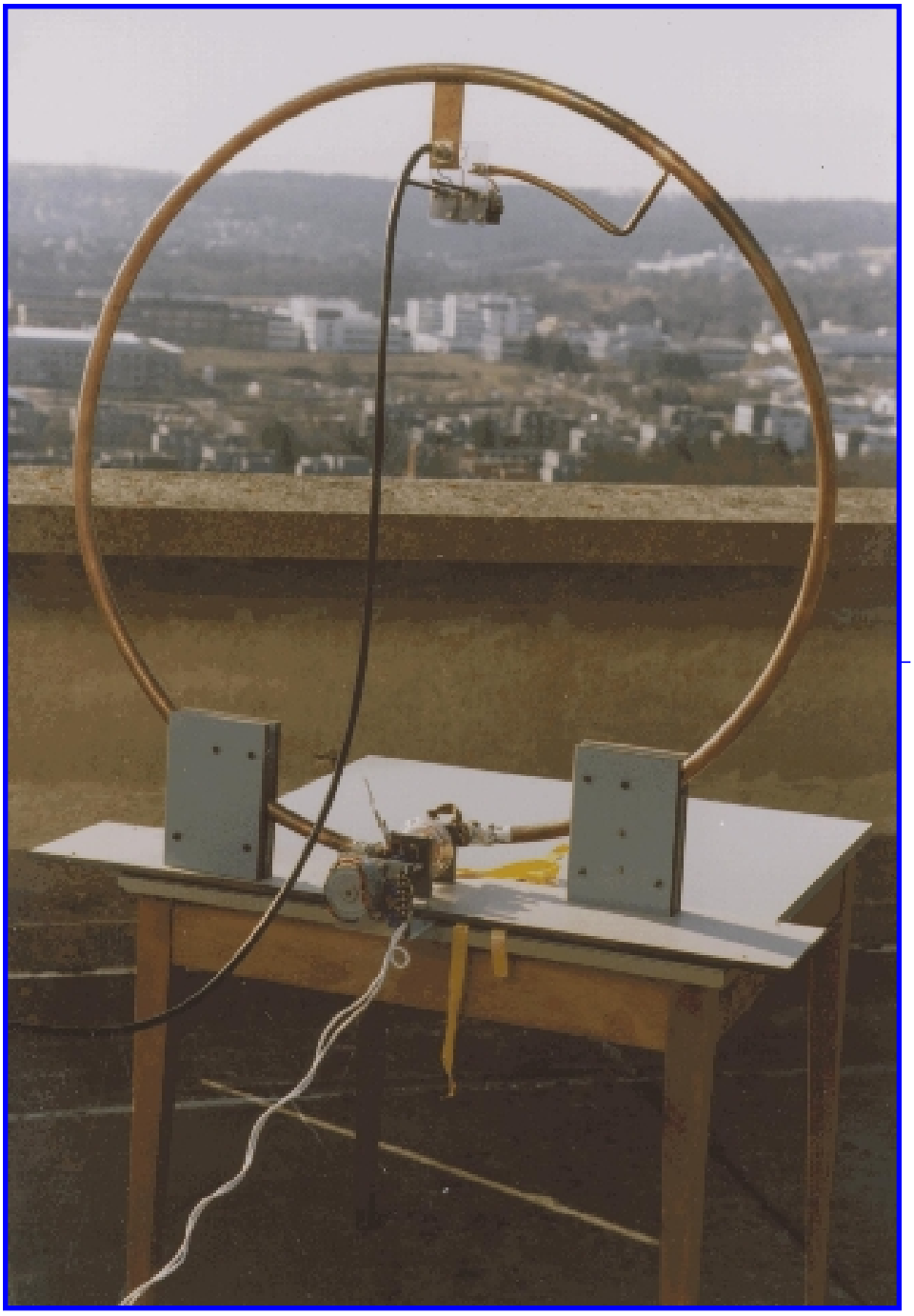
Coming Soon

More Projects, A Motor Controller Page, and More Articles from Other Loop Builders

Please email me at brian@standpipe.com

Copyright 2002 Brian Levy

Wellcome on DJ3TZ's Small Tuned Loop Antenna Page !



My first homemade STL ``Mag-1".

(Mni tnx to Markus DL5OBZ for the photo! Click for a larger version (719kB).)

I intend providing more detailled information here, but so far there are only some links.

Links to other ham radio small tuned loop (STL) pages:

- Guenther DH7BZ gives details about his homemade STL and practical experience (in german)
- Visit the homepage of Carl GW0TQM with plenty of information on almost every topic related to STLs, including many references and links. (page still active?)
- A collection of homebrew antenna links maintained by Rod AC6V, with many pointers to STL pages.
- An article archive also covering loop antennas, maintained by Willian Eric WD8RIF, including STLs.
- Peter VK3YE gives a detailled description for a homemade STL
- Stephen AA5BT describes STL construction
- Tony ON4CEQ describes his homemade STLs. Don't miss the photos!
- John G3PTO describes the G3BGR Magnetic Loop, lots of antenna links
- Jindra OK1FOU gives a detailed description of his homemade STL
- Harry SM0VPO/G4VVJ describes his 160m/80m loop and a remote antenna tuner
- Frank G3YCC describes the STL published by Tom GM3MXN and the "Rock Loop"
- Jess Dypin N0TFI also presents information about the "Rock Loop" (page still active?)
- Don't miss the "Loop Antenna Discussion" article archive
- Minowa 7N3WVM describes a STL made from printed ciruit borads
- David PA3HBB/G0BZF describes practical experience with building and using loop antennas
- Folkert PA3CQR gives information about his STL covering 10 - 17m

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Capacitors:

- Click here for references to articles about homemade HV variable capacitors in amateur radio magazines
- Tony ON4CEQ describes in detail how to make your own air variable butterfly capacitor
- RF Parts offers high power air variable capacitors
- Jennings is well-known for their vacuum variable capacitors

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There are several programs to compute STLs:

- FTP site with lots of ham software, includes STL programs
- Many ham-related computer programs written by Reg G4FGQ, including STL computation

---

No time to build your own loop? Click here for links to STL manufacturers and reviews of their products:

- Small Tuned Loops by WiMo Elektronik
- AMA (german abbreviation for "tunable magnetic antenna") loop antennas made by Kaeferlein electronics
  Click here for references to reviews in amateur radio magazines
- The High-Q Loop Antennas by MFJ
  (Their pages also contain the manuals.)
  Read the experience of Bob VO1DRB/WA6ERB/VE2DRB using this antenna for QRP operation
  Click here for references to reviews in amateur radio magazines
- AEA seems to be off market, but you might find their Isoloop Antenna on a flea market or so.
  Experience report ``Work the World with an AEA IsoLoop'' by Don W6TNS
  Click here for references to reviews in amateur radio magazines
- Broadcast band RX loop by Kiwa Electronics

Did any link change? Are there any good pages missing? Let me know!

Any comments? Click here to read and sign my guestbook

Please do not bookmark this page but only my start page http://www.qsl.net/dj3tz/index.html as the structure of my pages might change.

Last modified: November 26, 1999

Back to DJ3TZ's homepage

Magnetic Loop Antenna = Mag-Loop, Small Transmitting Loop

Felix Meyer

(Update: March 8, 2003, Dec. 17.03, Jan 5,04, Jul. 4,04)

The magnetic loop antenna is an extremly efficient short wave antenna for the small size it constitutes. It consists of a loop radiator made of copper or aluminium tubing and a tunable capacitor.
The size of the antenna is very small as compared to the size of a traditional antenna as dipole, beam, quad or vertical. The diameter of the loop is in the range of 1/10 to 1/100 of the wavelength.
The antenna works with the magnetic component of the EM field, which extends to the both EM components on larger distance. For that reason the antenna operates well close to ground and radiates a much stronger

signal than a dipole when both are close to ground. Surely, a full size dipole mounted in its optimal hight radiates better than a magnetic loop, but due its efficiency at low height the magnetic loop is an excellent portable antenna or may be used well as indoor antenne when external antennas are not permitted.
The capacitor of the antenna needs to be remotely adjustable to allow a frequency tuning range from 1:2 or 1:3. When properly built, the SWR is below 1.1 on the tuned frequency over the full tuning range.
The bandwidth is always very small and covers only a few KHz. The high Q of the antenna allows a selective receiption and suppresses effectively QRM of nearby BC stations, as well as other QRM.
Here I am using 2 loop antennas, one for 3.5 to 10 MHz and one for
14 to 30 Mhz, both antennas with only 85cm diameter, below the roof.
The 14 to 30 MHZ antenna with 50 W output allows for regular worldwide contacts with good results.
The antenna can be built easily as homebrew project if one can find or build a suitable capacitor.

Below you will find a loop calculation program for your own design together with detailed instrucions for magnetic loop antennas from 3.5 to 30 MHZ.

Calculation program LOOPABXE.EXE click for free download (new version Dec.31, 2003) with
calculation of multiturn loop, calculating all electric data.

>>> Explanation of program changes here.

For those who are interested: Formulas used by HB9ABX to design mag. loop

Circuit for motor control **Diagram 6th. Aug 1999**

Construction of Mag-Loop **(portable / indoor 14 - 30 MHZ)**

Picture of loop 14 - 30 MHZ **(Photograph thanks HB9DRJ)**

Construction of Mag Loop 3.5 - 10 MHZ

Picture of Loop 3.5 - 10 MHZ **(Photopgraph thanks HB9DRJ)**

Control through Coax

Magnetic Loop Antenna : Construction hints

The following instructions should be observed
for successful construction and operation of magnetic loop antennas:

DANGER : IMPORTANT NOTES !

The radiated field is very concentrated and may produce health problems.
Therefore, one has to keep distance to antenna of at least 5 meters
if the power exceeds 10 watts.

Coupling to the loop is done mostly at the lower side of the loop and the
tuning capacitor is placed on top. Due to mechanical stability I installed the capacitor with motor on the bottom and the coupling loop on the top without having any deficiency in HF radiation.

There are several coupling loop constructions. The simplest one is by forming a loop of installation wire (bare or isolated), or using a coax cable (type RG58). The diameter of the coupling loop is about 20% of the diameter of the transmitting loop. One end of the coupling loop is soldered to the center conductor of the coax, and the other end of the coupling loop is soldered to the braid of the coax.
There are several types of feeding: either capacitive coupling or inductive coupling. The easiest one being just a simple wire loop as you see here.
Here you see a capacitive coupling and here you see a gamma coupling.
The position of the coupling loop is inside the loop, exactly opposite to the capacitor. At this point, the voltage of the transmit loop is zero.
The distance between coupling loop and transmit loop varies from 0 to 6 cm.
By changing this distance, the lowest SWR is adjusted. Fine adjustment can be done by changing the form of the coupling loop, wider or smaller.

A good coupling loop is the symmetrical loop which you see on this picture.

The environment close to the loop influences this adjustment.
With proper adjustment a SWR lower than 1.1 can be reached.

Main loop and coupling loop should not be connected directly, as RF coupling to the feeding coax may appear easily and can produce RF interference (TVI/BCI), therefore no direct connection between center of main loop and coupling loop.
However, the main loop may be charged by static electricity and discharges by producing QRM bursts. This can be eliminated by inserting a small coil between the center of the main loop and the coax braid of the feeding coax.
If this connection is done without a coil, radio interference (RFI) may be

**produced due to small misbalance.
Coil data: 40 turns of .2 to 0.5 mm enameld copper wire at 1cm diameter
wound over 2 cm length.**

**The main loop may be made of tubing (copper or aluminum) or thick coax cable. If coax cable is used (RG213 or RG8 or similar) the inner conductor and the braid (= shielding/ground) is soldered together at both ends. These ends are then connected to the capacitor.
Very high current flows in the main loop. Therefore, thick and short copper wires are required to connect the capacitor.**

**In tubes, the current flows only on the surface due to the skin effect,
therefore the use of foil is an interesting method.
Very efficient and lightwight loops can be built by using a thick plastic
support and placing copper foils over this structure. The foil can be placed
in narrow strips in the direction of the loop circumference, as placing
the foil in one step produces crumpling.**

**The form of the main loop may be square, n-square or round.
The round form is most efficient as the losses are minimal.
(Best ratio of L:R).**

**Nearby environment affects the SWR.
In free field, the body of the loop should be 2 loop diameters above ground.
Good are 5 loops diameter hight, higher elevation gives only small signal difference.**

**If the loop is installed below the roof, then keep 10 to 30 cm space below
roof brick and avoid nearby lines and metallic constructions.
Note that the roof above the loop should not be a closed metallic construction.**

It's important to observe that the feeding coax below the loop is kept

**in the symmetric center between the two half loops straight downward at the length of one loop diameter.
By not observing this rule, RFI may be generated !**

**In any case, I recommend to insert a broadband current choke into the feeding coax, about 1 m away from the transmitting loop, as surface currents
on the coax cannot be prevented completely.**

**The loop capacitor needs to withstand high voltages and high currents. Butterfly capacitors are a very good choice as they have no sliding contacts. Variable vacuum capacitors are an excellent choice.
100 watts RF power produce about 4000 volts on the capacitor (see program). Required distance between plates is ~ 1 mm per 1000 V.
Suitable capacitors are made by HB9TJX (address at end of page).**

**A DC motor with strong reduction (about 2000:1) serves to control the capacitor. Suitable motors can be found in airplane or car model shops.
(E.G: Robbe No.4103 with 2430:1 reduction).
Grill motors may serve also fine.
Please note that there is a high voltage isolation required between motor and capacitor.**

I recommend to use pulsed DC current to control precisely DC motors. By adjusting pulse ratio properly, small increments may be controlled perfectly.

**A suitable circuit diagram is included for that purpose.
(Switch: FAST/SLOW; keys: RUN und BACKWARD).**

**The remote control motor may be fed through the same coax feeding the loop, hereby no separate control cable is required.
See link under: Control through coax.**

**Butterfly capacitors of different values for 6 KV are available at:
Markus Reber, HB9TJX
Eystrassse 7, CH-3400 Burgdorf, Switzerland, Tel: ++41 (0)34 423 16 79**

See datasheet at http://www.qsl.net/hb9tjx

Vendors of Ferrite Toroids:

- http://www.oselectronics.com
- http://www.palomar-engineers.com
- http://www.cwsbytemark.com

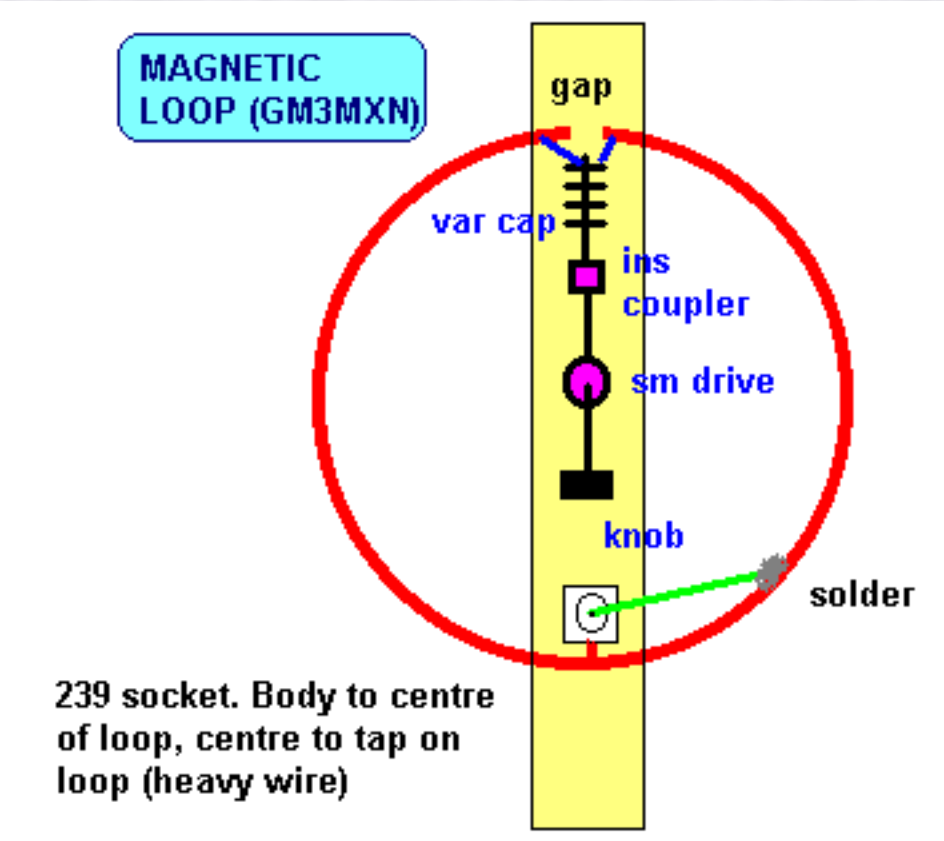
Please send comments or questions by e-mail to: see here (click)
or by Packet Radio to: hb9abx@hb9eas.che.eu

Manufacturers of magnetic Loops:

- **MFJ Loops** (good experience, cheap) (Product Catalog, Antennas, MFJ-1786, MFJ-1788)
- Christian Käferlein, Darmstadt, Germany, Tel:(+49) 6151 61272, Fax (+49) 6151 663009
  Kaeferlein DK9CZ (good, but not very cheap ...)
- **HB9CRU**(I3VHF Magnetic Loop Antenna)

Back (HOME)

A Magnetic Loop Antenna



Various articles have appeared describing magnetic loops for HF use, some with complicated methods of remote tuning and the antenna can be quite a useful and interesting project.

The article in Sprat 61 my Tom, GM3MXN describes a loop useable from 7 - 21 mHz, using half inch copper tube 3 feet in circumference. As shown, there is a gap at the top of the loop, which may be about 3/4 inch, either side of which is connected to the tuning capacitor, which can be about 250 pF. The inner of the SO239 socket at the base of the loop is soldered onto the centre of the loop, with the inner forming a gamma match to the tube, about 9 inches from the earthy connection. This is adjusted for minimum SWR. As show, the loop can be mounted on a wooden support and is fed with 50 ohm coax.

Notes

Tuning of the capacitor is very critical, and it may be advantageous to use more than one slow motion drive, which is what I did in my version.

It should be noted that a high potential is present across the gap, and even at QRP levels can be dangerous. Certainly with higher powers, this situation can be **LETHAL**. I would personally not recommend the use of such a loop in the shack because in my opinion even, QRP levels can produce what may be dangerous amounts of RF, which has been blamed for brain damage. Suffice it to say that I

know of one local who did use a mag loop indoors and wondered why the paint next to the loop was bubbling (!), and another who used a handheld VHF/UHF handheld, always with the rig next to his head. He died of a brain tumour. Whether there is any medical evidence or not, I think it wise to avoid situations of RF close to the body. *Is it worth the risk?*

For high power use a wide spaced capacitor is needed to avoid flash over.

Maximum radiation is off the ends of the loop, not broadside on, as in a quad or delta loop.

It will be necessary to retune the variable capacitor when QSYing, even slightly, as this is a high Q system. I used one on 40 meters and it proved fairly directional and quiet.

Other versions have used coax inner instead of copper tube. It is important to ensure that the loop is made from low ohmic material. If a square version is attempted, using say plumbers' connectors, ensure each joint is very well made, electrically.

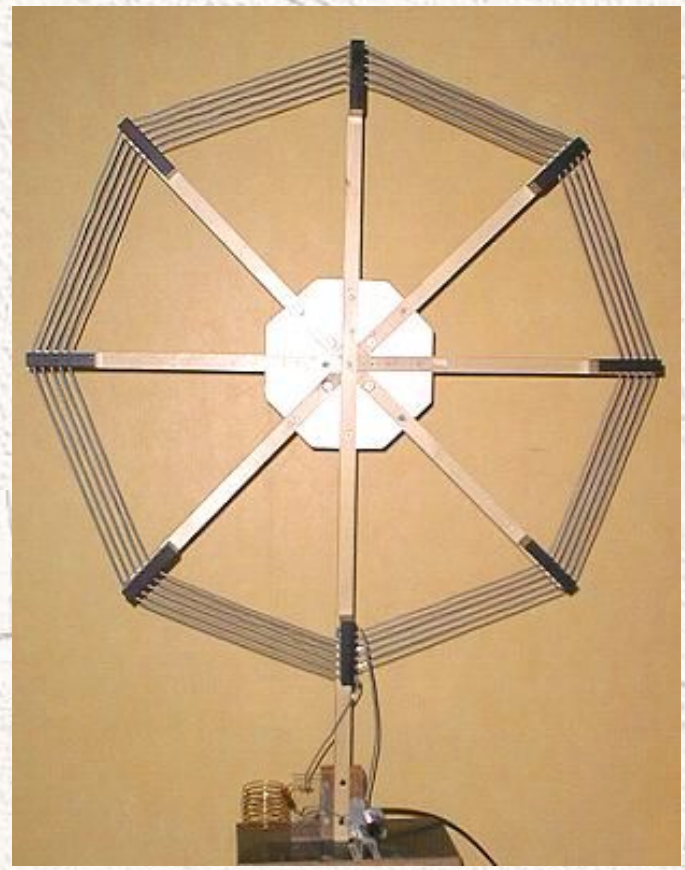
Click here to return to first page

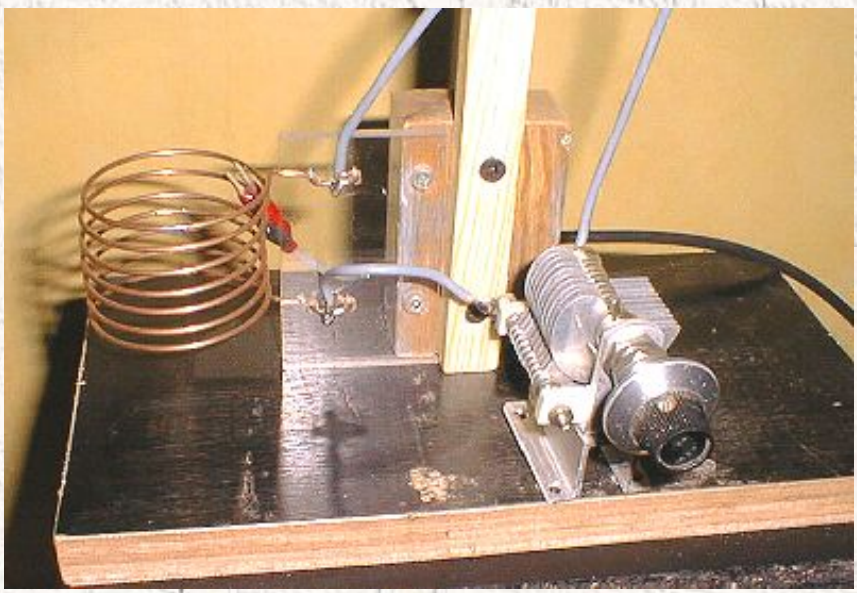
Magnetic Loop Antenna's.

Why is an Magnetic Loop antenna so special, this antenna is picking only the MAGNETIC part of the ELEKTRO MAGNETIC radio wave. The big advantage of this antenna is that the electric interference from the big city (streetlights, television's , cars etc...) have no influence on the received signal. With the loop you can hear other stations that you can't hear if you use a DIPOLE, with a dipole the stations are buried in the noise.

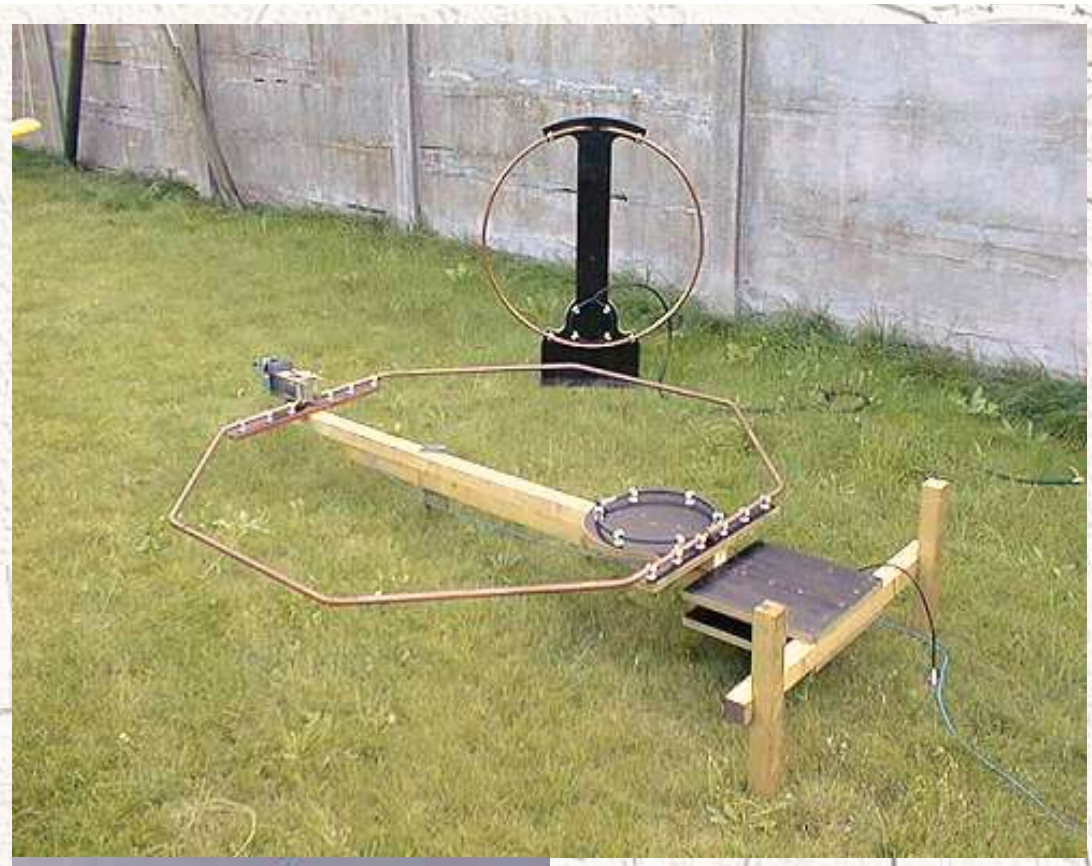
Multi Turn Magnetic Loop.

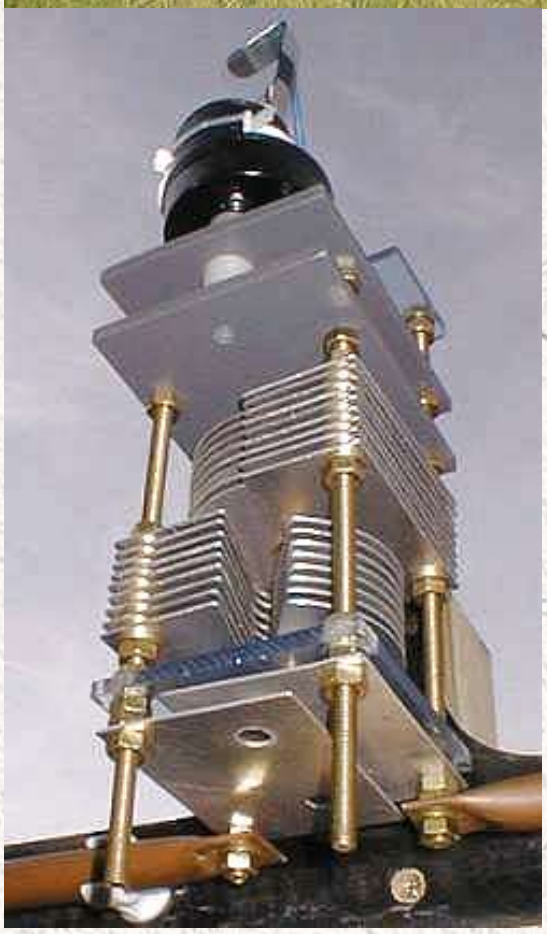
This is the first loop I build from a article in the QST from February 1996, it's 30 Inch-diameter, and it's designed by G2BZQ/WØ for 80 M.





Single Turn Magnetic loops





The first one turn loop that I built was made from 75 Ohm TV Coax and with a small explanation in the RSGB handbook for radio amateurs. I used the outer screen from the

coax and the results of the loop where good.The next loop I built is a octagon loop in 15mm copper tube with a circumference of
4.8 meter (16 feet).
The frequency range of this loop goes from 14 MHz to 7 MHz and works fine. The biggest problem is the tuning capacitor, if you transmit with a power
of 100 W you need a capacitor with a voltage rating of 5000 Volt.
A capacitor which can handle this voltage is hard to find over here and if you find one they are very expensive.
The first capacitor I built was a design from GW3JPT from a article in the RADIO COMMUNICATION from February 1994. It is a split stator capacitor with a capacitance of 140 pF and with a voltage rating of 6000 Volt.
The capacitor is remote tuned with the use of a small BBQ spit motor.

The second capacitor I built is my own design and it's a butterfly capacitor because the losses are lower than a split stator.
The capacitance is 5-65 pF and the voltage rating is 7200 volts. I used
it for the small loop with a dia. of 800 mm (2.66 feet) and the frequency range of this loop is from 28 MHz to 14 MHz. The Aluminum plates of 1 mm
for the capacitors are cut with a JIG SAW.

Most asked Questions:

I`d like to talk a little more on your setup. it seems like something whichI could get together if only some more data was available. do you have anynotes etc still laying about since its build ?

The theory for calculating the loop is very simple.The circumference of a magnetic loop is 1/4 wave of the designfrequency.

Example for 14 MHz.

300 / 14 MHz = 21.428 m is 1 wave

21.428 / 4 = 5.357 m is 1/4 wave circumference

5.357 / 3.14 = 1.706 m diameter.

The recommendations are that you can tune the loop from the design frequency to the frequency divided by 2 to keep the effiecency acceptable.

14 MHz / 2 = 7 MHz

I made the small loop (800mm / 31.5 ") from soft copper tube on a role that you can buy in a plumbershop and it's easy to make a nice circle if you draw on the ground a circle with a rope and a piece of chalk.

For mounting the loop to the hardboard I used plastic clamps
that they use for mounting copper tube on the wall.

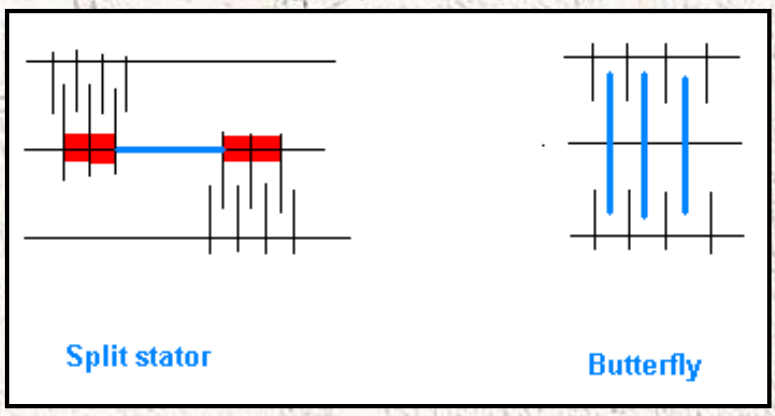
---

why is a butterfly capacitor better?

For high voltages and currents the use of Capacitors with wiper contacs is not recomended. That's why they use capacitors in serie's.The pro for serie's capacitors is that the voltage rating is doubled.The anti is that the value of capacitance is divided by 2.

For the split stator capacitor the 2 capacitors are connected in series by the shaft (bleu) and the red spots on the first drawing are losses.

For the butterfly capacitor the 2 capacitors are directly connected in series by the rotors and gives less losses.



Do you know of anyone that has built a similar loop that outperformed a garden variety dipole?

Compare antenna's is very difficult , sometimes I have for 60 % better signals in RX and TX on the loops then on the dipole.

In Theorie is the performance of a magnetic loop - 0.4 dB lower then a dipole or a vertical .

I have over here a homebrew trap dipole from 40-20-15-10m and the height aboveground is only 7 m(23 ft), for a good performance on 40 m the dipolemust have a height of 1/2 wave above ground (66ft). I don't have a radiation angle on 40 m and it's only good for contacts in Europe and not good for DX, now the 1.5 m loop tuned to 7 MHz with a effiecency of 38 % (38 w ERP) and a angle radiation of about 20 degrees performs better than the dipole because the vertical magnetic loop only 1 M above ground as a angle radiation and the dipole don't.

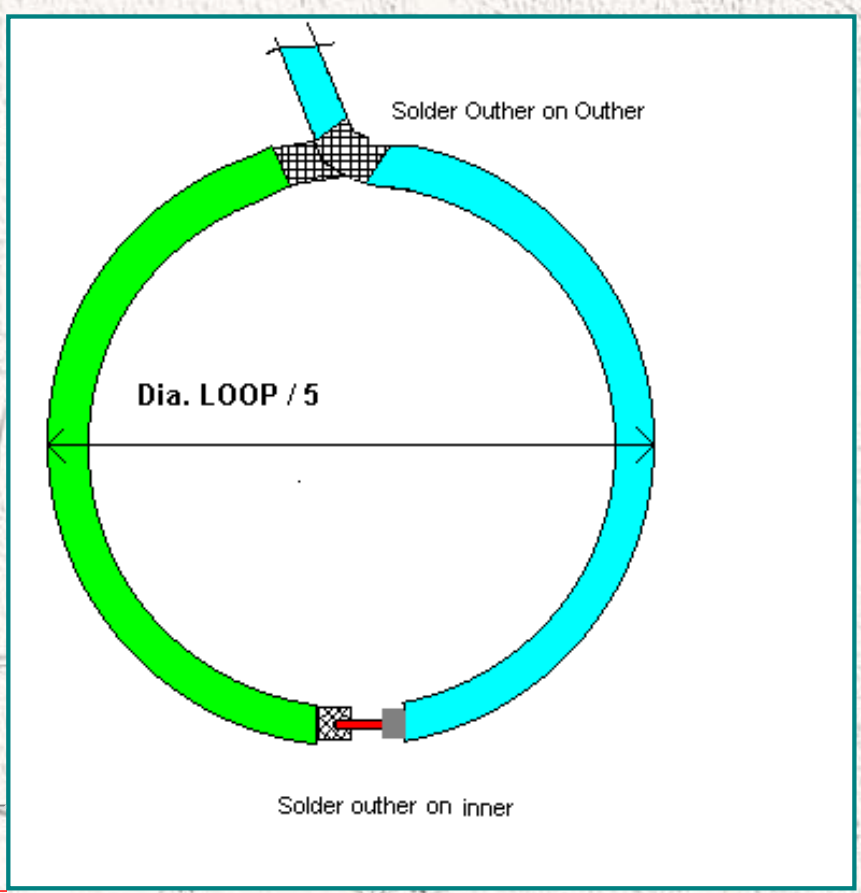
Another advantage is that the reception on a loop is mutch better, on 20m I have with the dipole S5 noise from the big city, if I switch to the loop I have S1 noise and hear stations who are buried in the noise when I use a dipole.

Coupling loop dimensions?

I find that the best way to feed the loop is with the shielded 1/5 Faraday loop made from coax RG213 or RG8, I tried the gamma match bud I had problems to keep the VSWR low on all Bands, the shielded loop gives on all bands VSWR 1.1 and reduce more noise pick-up then the gamma match.

I found out that if you use a 1/5 Faraday loop, that the loop is to big, making the loop smaller with 0.5 inch by the time in circumference and checking with a field strength meter you can see that the radiated power increase.

The place off the feeding loop is placed at the electrically neutral point on the loop and that is 180° from the capacitor and I have the best results with the feeding loop close to the ground and the capacitor far from the ground.



I was wondering if you worried about the resistance of the mechanical joints (copper pipe bolted to the capacitor) significantly reducing your radiation efficiency as I think the radiation of these antennas can get as low as .01 ohms

Soulder or weld the capacitor plates is always the best, but I'm afraid if you make the spacers and the plates in ALU that with welding everything is gona bend from the heat

and I know from practice (I work in a maintenance workshop) that welding ALU is coarse. Another possibility is using all brass or copper and solder, there are hams that using double PC board for the plates.
I made a QSO in phone with Florida, RPRT 5-5 and the other station used a vertical antenna, with the small loop (800mm and theoretical effiecency 41 % on 14 Mc) vertical in the garden and the states side is through the house. I was very happy with the results , so I think that a capacitor
maded with torqued compressed joints is good enough for using 100 W.

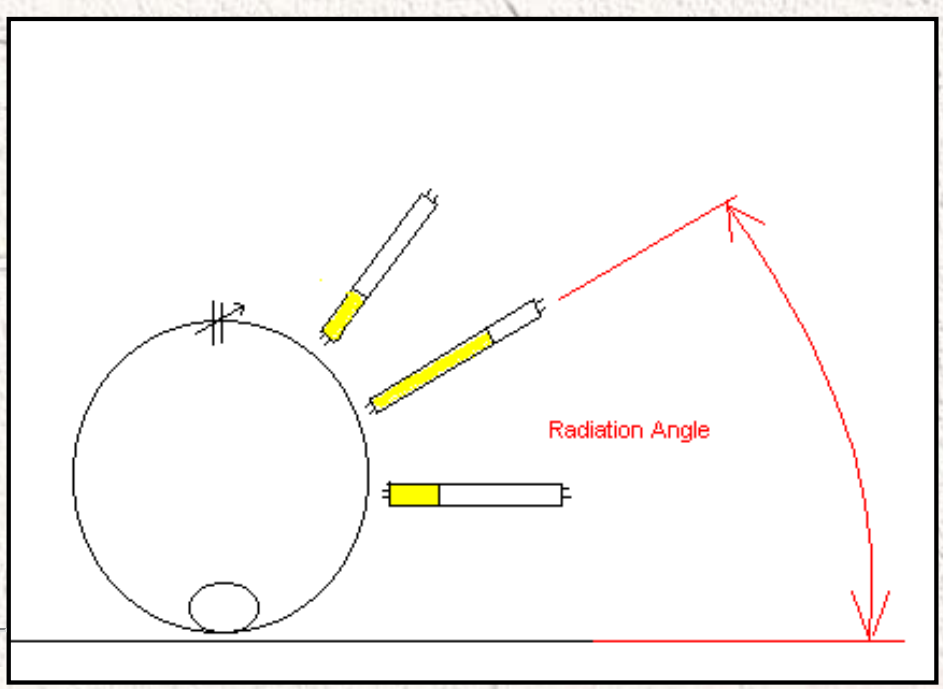
Have the dissimilar metal joints weather well?

To keep the oxidation low on the dissimilar metals I used a thin coat of vaseline after assemble the capacitor and with the tupper ware a like plastic box it is good protect against all wheather conditions.

How to find the radiation angle of the antenna?

Can it be found practically?

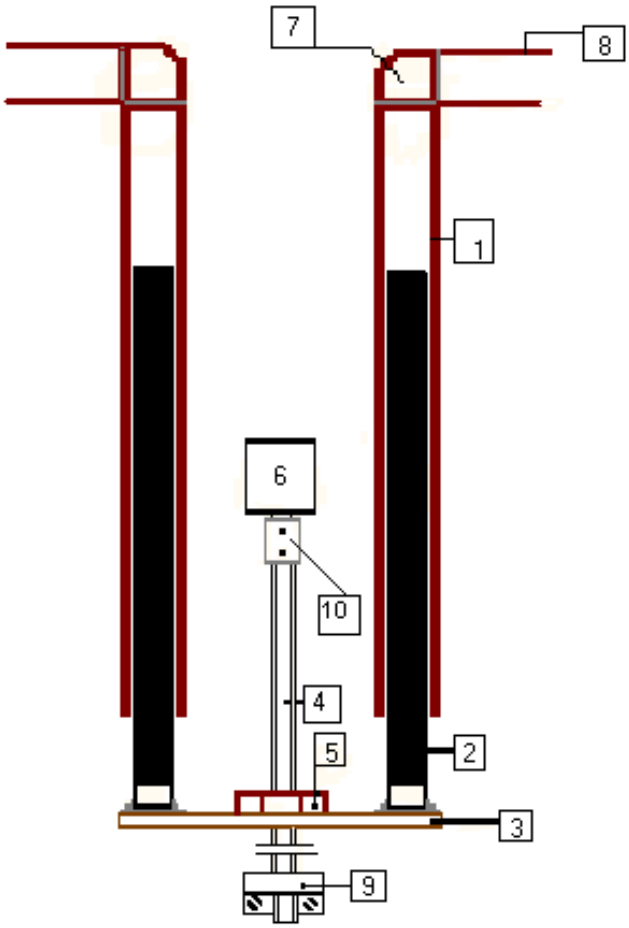
Finding the radiaton of a magnetic loop is very easy, with a TL-lichgt tube you can see it, with abt 10 w power on the the loop with the TL-tube in the plain of the loop at right angle to the circle you see the tube lightning, there where the the light is the farest on the tube thats the radiation angle.



When you refer to washers, nuts and rods you use the term "M6".Please forgive my ignorance, but to what does "M6" refer? Does this mean 6mm?

M6 is (M=metrical) and 6 is indeed 6 mm threaded rod and
you can compare the size with W1/4" (6.35mm) .

A very easy to build Piston Capacitor.



1) Copper tube

2) Coax RG8 or 213

3) Double PCB

4) Threaded Rod 1/4"

5) Brass nut

6) Motor with reduktion or stepper motor

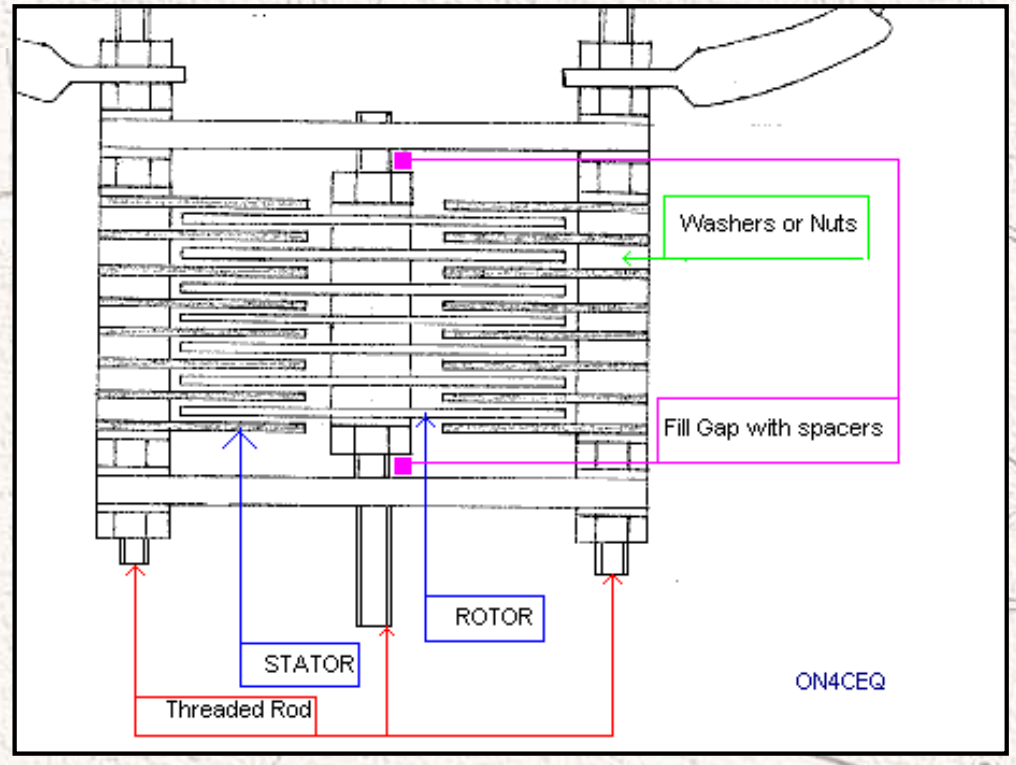
7) Copper Elbow 90°

8) Loop

9) Bushing from old var resitor

10) Couple shaft (PVC)

How to build your own Butterfly Capacitor.



The best material for the front and the back is CLEAR PVC 3 or 5 mm thick as alternative you can use GREY PVC or 2 sheets pcboard together with the copper removed .

The best material for the washers, nuts (M6) and threaded rod (M6) is brass or stainless steel,(NON MAGNETIC MATERIALS for the losses).

For the spacing of the vanes you can use 2 washers M6= (6Kv) or a nut M6 =(12 Kv) if you use aluminum plate 1 mm thick.

If you use a nut then the best thing to do is remove the thread by drilling withØ 6.2 mm.
The effective area for the vanes is 11.7 cm² and with the formula

for 2 washers = (0.0885 x 11.7 cm²)/ 0.1 cm = 10.35 pF for 1 air gap.

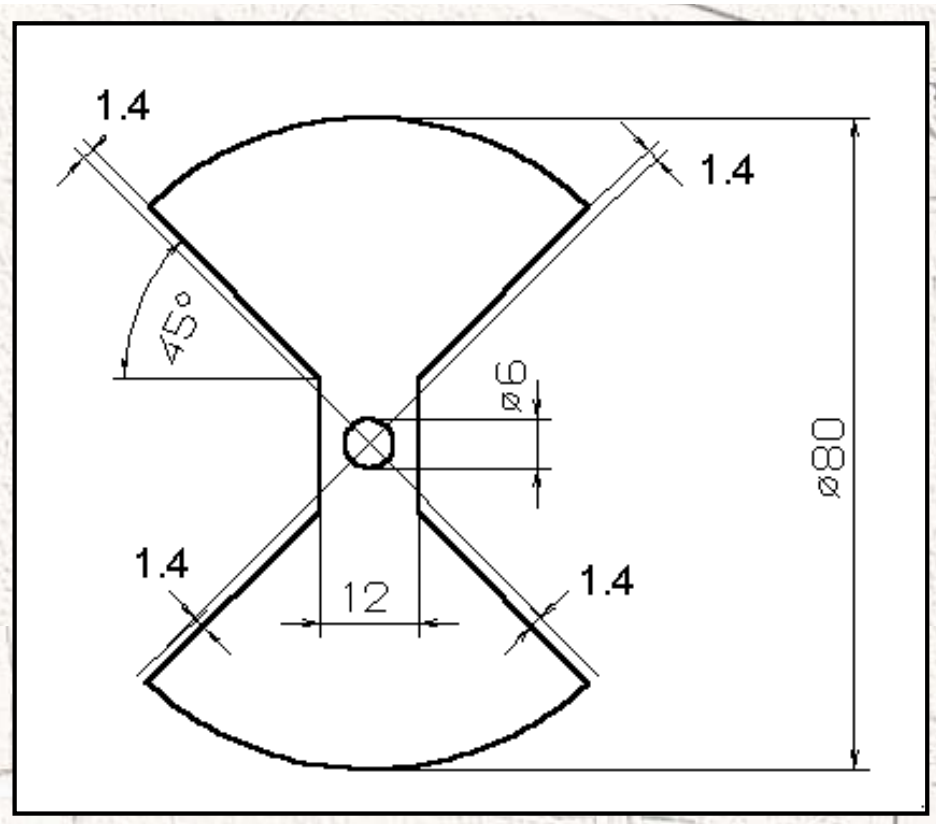
for 1 nut = (0.0885 x 11.7 cm²)/0.2 cm = 5.17 pF for 1 air gap

Example:

If you you make a capacitor with 2 washers as spacing and you make 5 rotor vanes and 6 stator vanes then you have 10 air gaps.

10.35 pF x 10 = 103 pF + 10 pF stray capacitance = 113 pF / 2 = 56 pF

The final result is a capacitor with a value from 5 - 56 pF.


1.4
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45°
ø6
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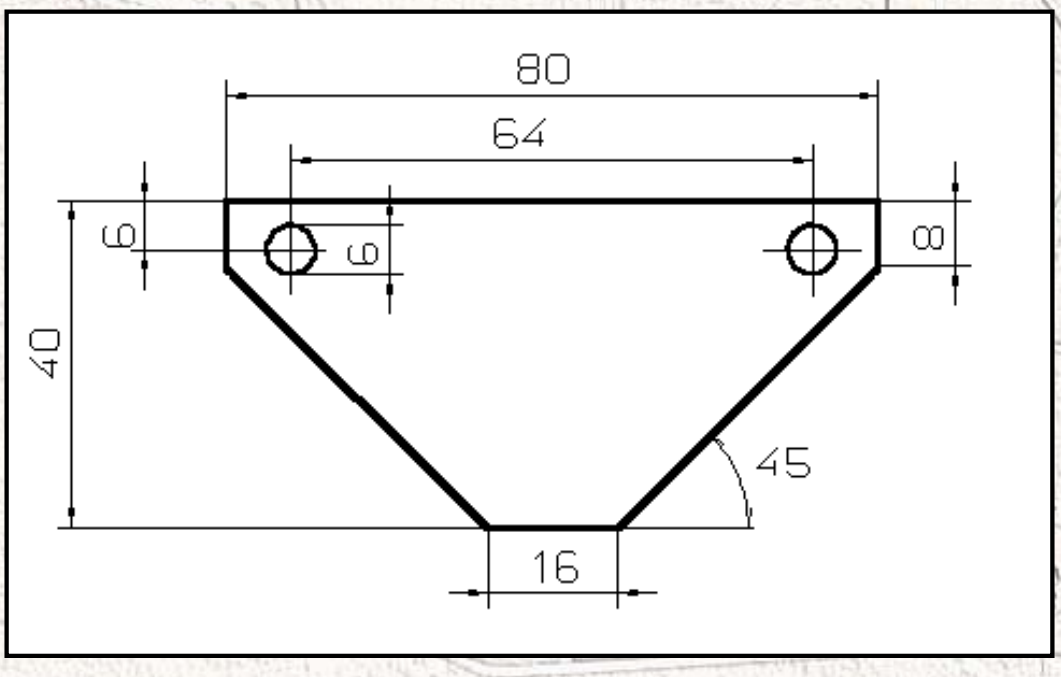
80
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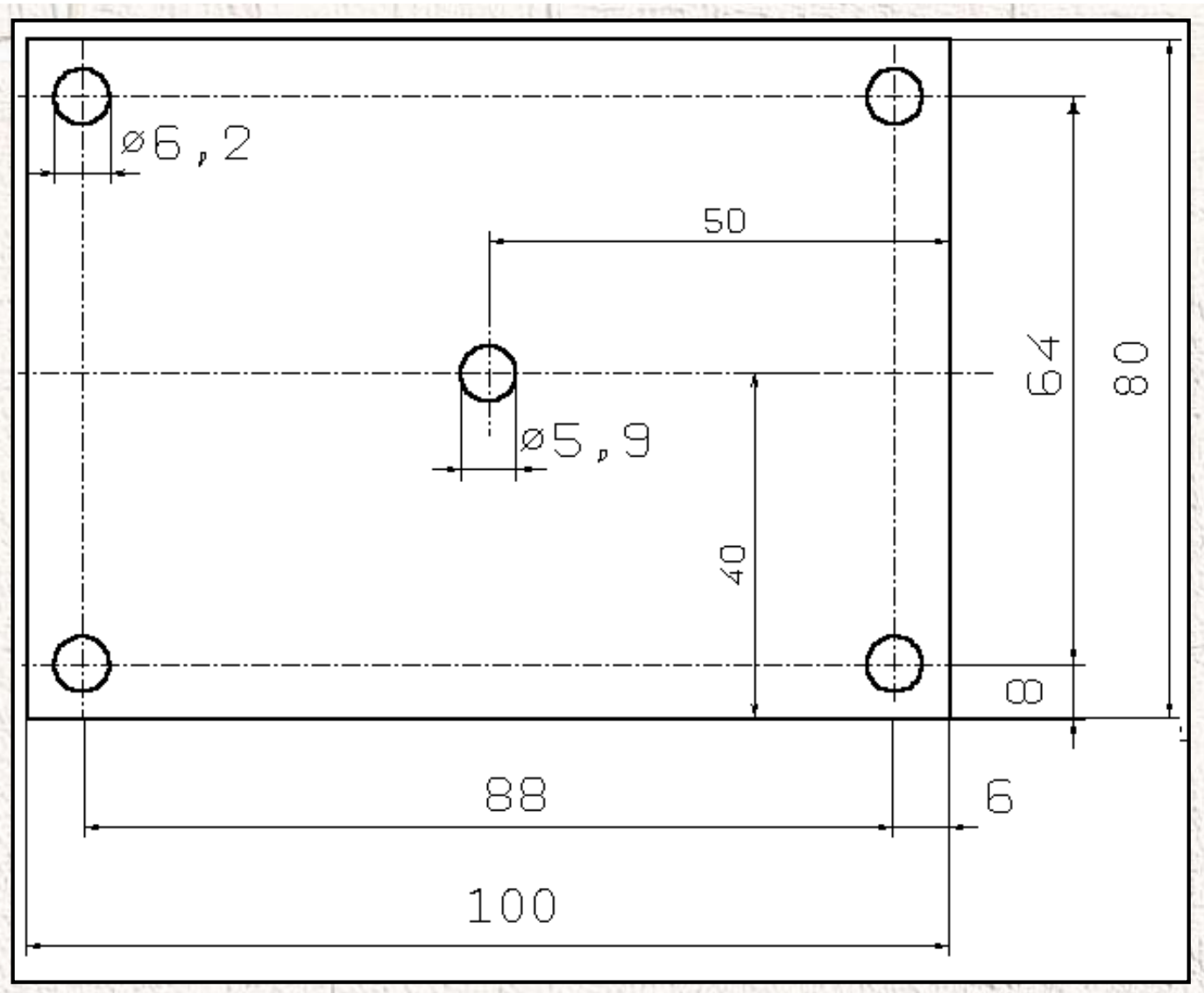
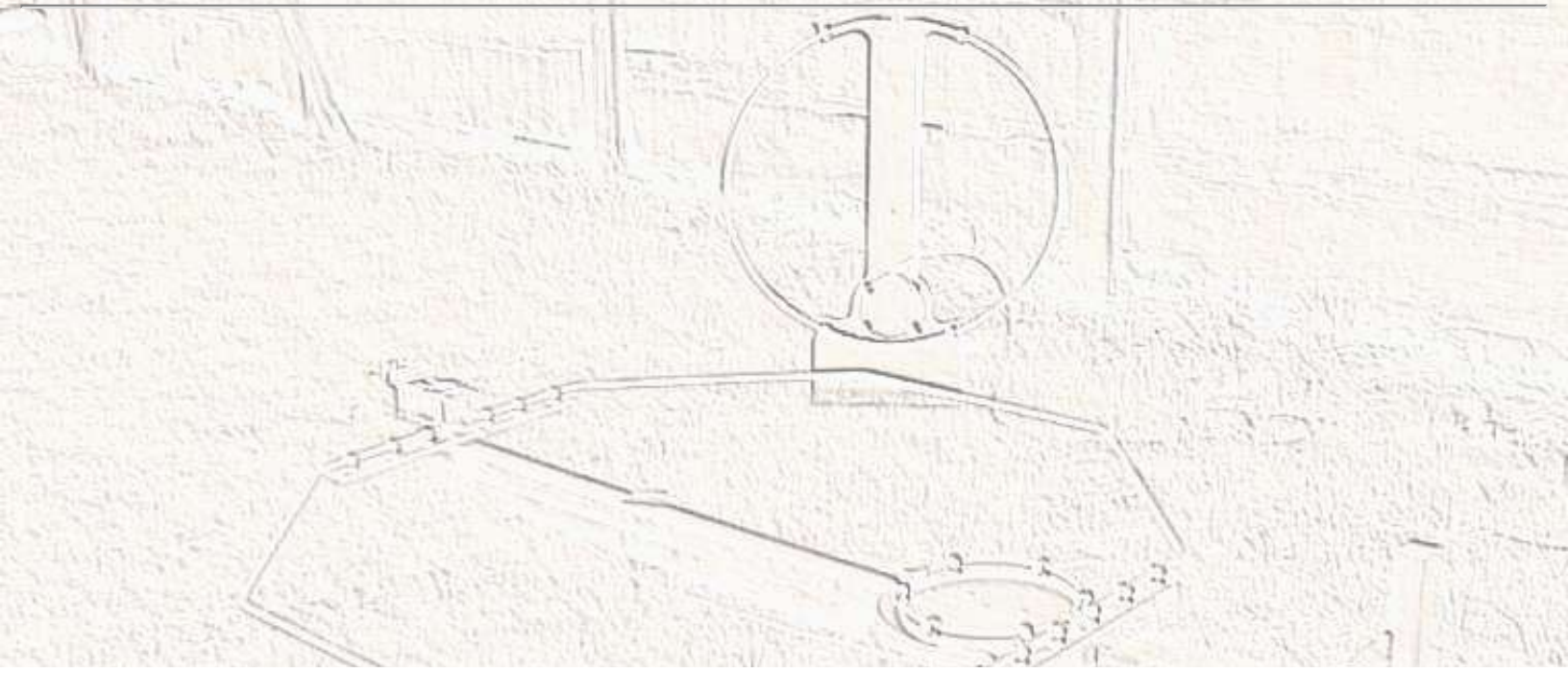
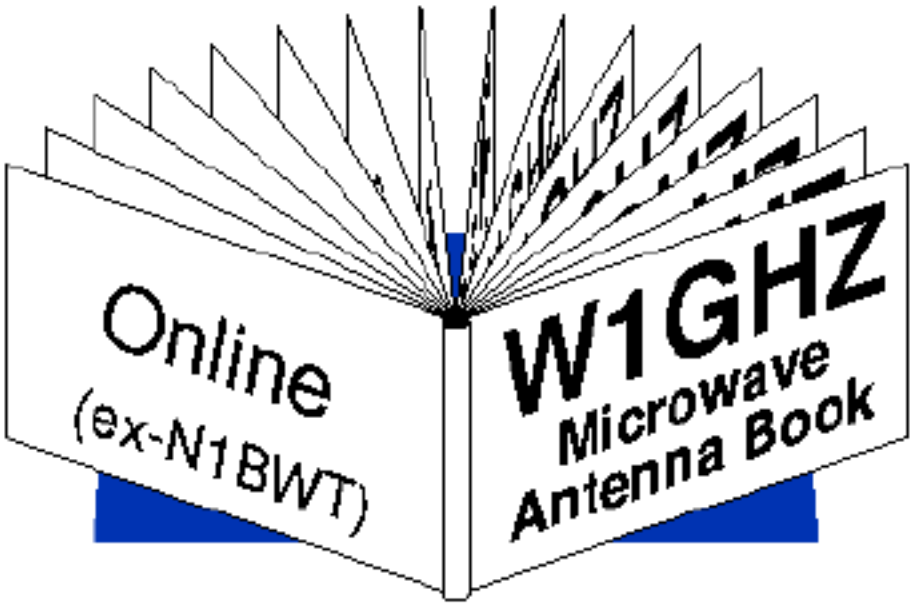


Last Update 03 December 1999. ON4CEQ Van Herck, Tony QTH: Anwerp(JO21FF) BELGIUM.





The W1GHZ Online Microwave Antenna Book

Paul Wade W1GHZ (ex N1BWT) ©
1994,1995,1996,1997,1998,1999,2000,2001,2002,2003

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Downloads:

HDL_ANT.exe *version 3b4* and Source code 3b4 *March 2003 - Minor corrections*
FEEDPATT.exe
PHASEPAT.exe
Periscopegain.xls
Slotantenna.xls *30 May 2002* ******Updated******

Copyright notice

Home Brew dual band antenna

Generally speaking, this is a dual band design focused with better performance on UHF. It can satisfy needs on most situation. If you always work in hot areas where out-band cross modulation are causing trouble to your reception, this antenna might even improve the selectivity slightly. If you always work at marginal signal area and have to squeeze the last drop of energy from your system, this is not a suitable candidate.

Basic design theory

This antenna is just a 1/4 wavelength resonator for both UHF and VHF band. It provides no GAIN as compared to other multi-section design. Yet it has the advantage of better stability against surrounding influence (e.g. different mounting method, unexpected reflection by surrounding objects, difference in stray capacitance between test bench and operating field). The antenna consists of two sections. The lower section is a conductor cut to resonant on UHF band. The upper section is a coil which serves two purposes. For UHF band, this coil is a RFC (radio frequency choke). It blocks UHF frequency energy to

flow through upper section. For VHF band, this coil enhances the inductance provided by lower section. The resultant inductor, together with surrounding stray capacitance will form a VHF resonant circuit and becomes the VHF antenna. To get a better understanding on the tune up procedure, it will be helpful to review some basic rules onantennas.

- Antennas are LC circuits. The LC components control their resonant (operating) frequency. Inductive component is formed by metal conductor of the antenna's body while capacitive component refers to stray capacitance between the antenna body and its surrounding grounded conductor.
- Increasing either L or C component can lower the resonant frequency while reducing them causes the opposite effect. Inductor can be increased by adding physical length of the antenna element or by making it as a coiled structure. Capacitance can be increased by using a larger surface area on the antenna body, by adding capacitor hat or by reducing physical clearance between inductive section from ground. In this issue, we mainly alter the inductive section of the antenna to achieve tuning. That is done by cutting length of the conductor and by changing length of coil structure.

Mechanical structure and material consideration

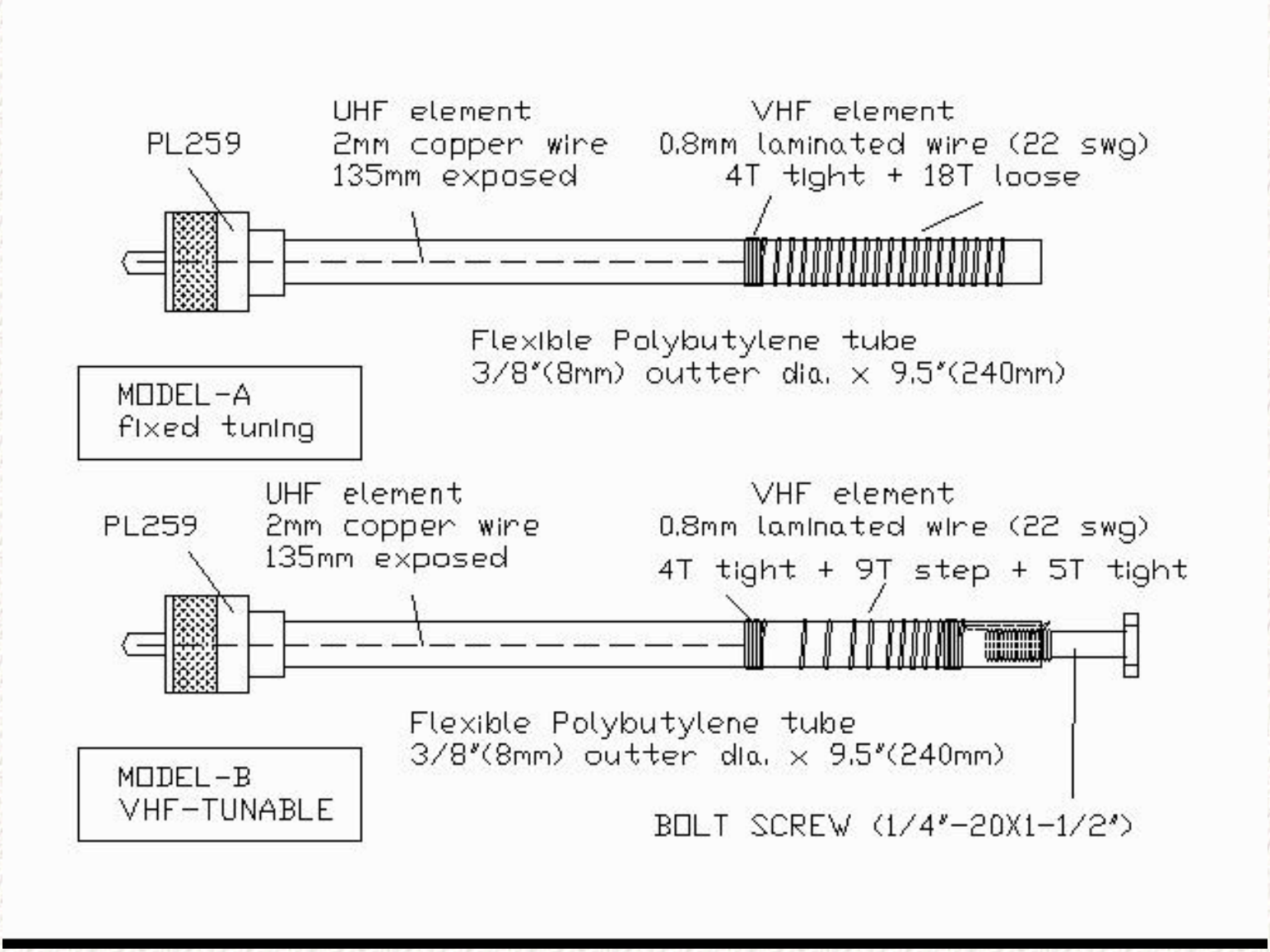
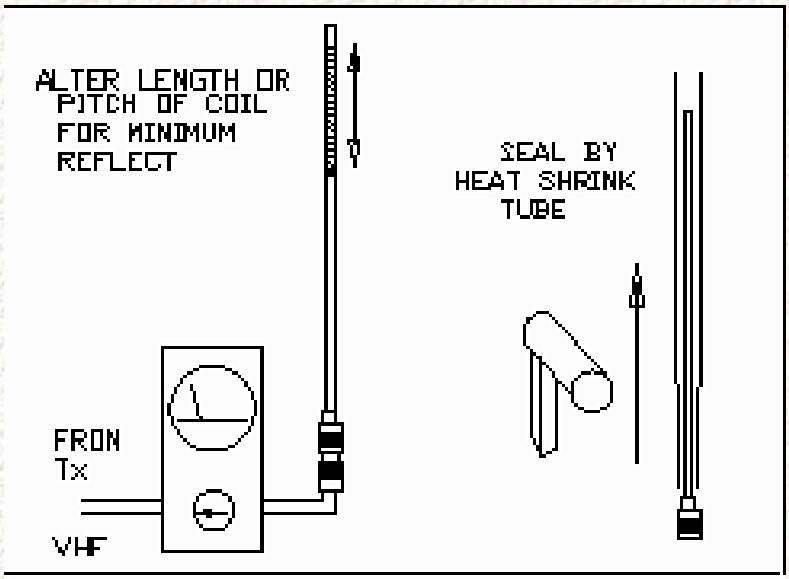


Fig.1 shows two possible structure of this antenna. Model(A) being the fixed style and model(B) being VHF adjustable. The lowest section of the antenna is a PL-259 plug. The center pin of which is soldered to a 2mm diameter copper rod. I got this copper rod from core conductor of RG-8U coaxial cable. It should create an exposed section (clear off from metal shield of the PL-259 plug) of about 135 mm. More detail will be discussed in tune-up section. This copper rod is hidden inside a tube which acts as the coil-form for the upper section. To avoid absorption of RF energy emitted from the antenna, material of this tube must be carefully selected. I used PVC faucet tube for this part. They are polybutylene - those gray color PVC plastic which is cheap and easily obtainable from most hardware stores. You can use any other tubes provided their external diameter can fit into the barrel of the PL-259 plug plus it does not absorb radio waves. We can use a microwave oven to check out if a certain material will absorb R.F. energy. Material placed inside an operating microwave oven will generate heat if they absorb radio waves. Place the material under test inside the microwave oven together with a cup of water (for safety). Execute heating process for say 1-2 minute and check surface temperature of the object under test. If the material stays cool as compared to the cup of water, we can assume that it do not favors radio wave absorption and will be suitable for the job. A small hole (1 mm) is drilled on the tube just besides tip of the UHF section. A piece of laminated copper wire (22 swg) is soldered to the tip of the UHF section and is leaded to outside of the tube through this hole. The rest of this wire is wound on the upper part of the tube and forms the VHF section. After proper tuning is completed, the whole antenna is sealed within a piece of heat shrink coating down to the neck of the PL-259 plug. A rubber cap can be placed on top of the antenna in order to prevent water from entering inside of tube.

Constructing procedure and tune up



Cut out a piece of 2 mm copper rod. Let's starts with a length of about 160 mm. That should cover future cutting job and the length hidden inside the PL-259 barrel which is not counted for radio emission. Insert it into the center pin of the PL-259 plug and make a smooth soldering at the tip of the pin. We now have a UHF antenna. Before attempting to tune it, prepare two pieces of 2 mm copper rods also, they should be of length 140 mm and 500 mm respectively. They will serve as the temporary ground planes for UHF

and VHF test. By using a reflectometer (SWR meter), check out the reflect power from this simple antenna when UHF power is fed. Cut the length of this section slowly (in 2-3 mm step) and observe changes on reflect power at lower end of target frequency band. Without attachment of a good ground plane, it is impossible to obtain zero reflect status so just go for the lowest possible value. Using the low end frequency for testing is a safety precaution. That prevents you from over cutting the antenna without notice. After you felt satisfied with the cutting progress and has slightly over cut the rod (reflect power increase again upon further cutting). Change test frequency to middle of operating band and proceed further cutting with high care. Once reflect power increased again (indicating slight over cut), pick up the earlier prepared 140 mm section and touch it to the PL-259 plug of this antenna. You should see the reflect immediately drops for a lower or even 1:1 value. This confirms that the section is properly cut and is ready. Cut out a piece of faucet tube (240 mm in length). Insert it into the barrel of the PL-259 plug. Make a mark on the tube where it just clear out from the PL-259 barrel. Remove the tube and put it side by side with the UHF antenna while the mark is aligned with the top of the PL-259 plug. Make another level mark for the tip position of the UHF section. Drill a 1 mm hole on the tube at that mark. Prepare for a piece of 1 meter long 22 swg laminated wire . Push it through the above mentioned hole and force it to run through the lower section of the tube until it made its way out from the bottom. Solder this wire to the tip of the UHF section and pull the wire back into the tube. That will bring the UHF section into the tube as well. Pull out any excess wire inside the tube until the tube is seated tightly inside the PL-259 barrel again. Wind the laminated wire on the tube for 22 turns. These turns should be in a pattern as shown in fig.1. The first 4 turns which forms the RFC for UHF band must be tight. The later 18T should be interleaved and evenly spread along until reaching end of the tube. That should make a coil of approximately 80-90 mm in length. Place the antenna on the reflectometer again, recheck the SWR or reflect status on UHF band, there should be no change as compared to earlier test result. Change frequency to lower end of target VHF band and repeat the test. Tune the VHF response of the antenna by altering length and pitch of the 18T section. Generally speaking, start with evenly distributed pitch and try to control the resonant frequency with coil length. If that cannot work out, try to rearrange the pitch unevenly - tighter pitch near the bottom and looser pitch near the top. Such treatment can further lower the operating frequency. Like tuning job in UHF band, go for the lowest possible reflect on the middle spot of your target band and confirm proper tuning by adding of VHF ground plane (500 mm long copper rod). Once both bands are confirmed in tune, seal and secure the whole antenna with heat shrink tube. The antenna is now ready for application at mobile vehicle environment. If you want future tuning possible in order to match different field environment, you can try model(B), the VHF tunable configuration. In that case, another 1 mm hole should be drilled on the tube near top portion of the coil. The 22 swg wire should be leaded back into the inner side of the tube. A section of contact wire 10 mm in length at top end of tube should be formed. A 1/4" metal bolt is screwed in to fill up the tube. Its pitch should bite through the laminated surface of that 22 swg wire and make a contact with the wire. The bolt will become an extension of the coil and by adjusting the bolt length (in/out or even changing length of bolt), the VHF resonant point can be adjusted.

Conversion for base station application

This antenna can be used at base station configuration also. The trick is to provide sufficient ground plane system in order to form those essential stray capacitance for proper tuning. Fig.3 shows a simple

dual-band ground-plane design and corresponding mounting for base station.

This mount can be used with other UHF/VHF mobile antenna design also (as long as they use PL-259 base connector). It is a very good test bench if you have needs to compare efficiency between two mobile antennas or just want to convert another mobile dual band antenna for base station application. The materials for this mount are all easily obtainable pipes and sockets used in electric power installation. We need a piece of electric conduit (metal pipe for protection of electric cable). Length is not critical, say 20" will be fine. It should be of the 0.5" diameter series. These pipe are measured for their inner diameter. Find also the EMT connector (set screw type) for the same family and put them on both ends of the conduit. The reason for using this dimension is because - the inner diameter of such EMT connector will perfectly fit a SO-239 socket (UHF female bulkhead receptacle solder type). You can tap threads on the inner wall of the EMT connector so that the SO-239 can be screwed in. I used a short cut, just hammer the SO-239 into the connector. The only weakness of this method is - the socket cannot be removed easily or frequently. Anything must be done right at the first move and cannot be repeated for correction. By putting two SO-239 socket on each end of the conduit and have them connected inside the pipe, a female to female adapter is formed. The upper socket is for mounting of the antenna, the lower one is for connection of feed line. The length of the conduit serves another purpose - made possible mounting of the whole antenna on another vertical pole easily with two hose-cramps. We also need a piece of 4" round-pan (top cover for those circular shape power junction box). It provides a flat platform where the ground-plane elements can be mounted. The ground-plane elements are made of threaded stainless steel rods (size 10-24). Holes at 60 degree separation are drilled on the vertical edge of this round-pan. The threaded rods can then be mounted through these holes with two nuts (inside and outside the round-pan's edge). Length of ground-plane elements control their resonance frequencies. Different materials needs different lengths to achieve resonance. Our earlier mentioned value 135mm at UHF and 500mm at VHF are only good for copper conductor. For steel, the length will be different. Togerther with situation that these rods are mounted 2" away from the RF socket, their dimension will be much shorter. Practicle test indicated that UHF ground elements in this design is best at length 70mm while VHF elements at 350mm. This length refers to rod's length between their tips and mounting point at round-pan. Since these rods are mounted with two nuts, their effective length can be adjusted. To play

safely, the rods should be cut with 30mm extra length. That comes out as 100mm for UHF and 380mm for VHF. Adjusting procedure on their final length will be discussed later. The SO-239 connectors has to be connected inside the conduit. I made a mistake at the beginning. Feeling that both sockets are tightly hammered and in good contact with the pipe, their ground points should have been connected nicely. I used only single core wire to connect their center pin rather then using coaxial cable to join them. But I forgot one thing, the two EMT connectors were secured on the metal pipe with only two set screws ! That is not a good contact for radio waves. I was surprised by the test result of such prototype - SWR reading of nearly 1.8:1 even with proper 50 ohm dummy load teminated. After modifying the connection back with standard RG-58AU cable, SWR for this section normalized to 1:1. Drill six holes on the edge of the 4" round-pan and punch out the center hole on the top surface. Mount the round-pan on the upper EMT connector, Secure it with the nut which comes with the connector. Cut threaded rods to proper length (three at 100mm, three at 380mm) and have them mounted on round-pan in alternate pattern. One UHF element followed by a VHF element then vice versa.

Construction of the mount is now almost finished (awaiting final tune up). It is not a bad idea to paint the whole thing - for protection against water damage in future. Try to do it after all ground-rods are set. We need good metal contact surface between the round-pan and the rods. Tuning of ground-rods are similar to tuning of main element. The longer the length, the lower the resonance frequency. Since we have three rods for each band, we can set one rod on the lower frequency limit, one on high limit and one for the center spot. This will create a flatter responce of the antenna in the entire operating spectrum. Always remember to work on one rod at a time (with the other five removed). Set up one test frequency and alter the rod's lenght for minimum reflected power. Then mark the length and change for another rod

with another test frequency. Put them back onto the round-pan with proper tested length after all key spots (three for VHF and three for UHF) are checked out. Reflection from near by objects can distort the above measuring result. These tests must therefore be done in open space, try to obtain clearance of at least 5 meter from any objects which might cause reflection. Always try to use high quality feed-lines to avoid creation of fake image of low reflection when reflected power are absorbed.

More detail on this antenna can be found in Radio Fun May 1994.

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MATTS SUB

4X4 STACKS 30 FT BOOMS ON 2.5 TON FREESTANDING ROTATING TELESCOPING GK HB SKYNEEDLE

THE 6 ELEMENT ANTENNA PARTY

ROTATING FULL SIZE 80 METER 2 ELEMENT QUAD FREESTANDING AT 125 FEET 40 FEET FRONT TO BACK SPACING

80 METER 2 EL ROTATING QUAD

MOBILE MOTORIZED HF ANTENNA

3-30 MHZ MOTORIZED MOBILE ANT

YAGIS STACKING

QUARTER-WAVE COAX MATCHING HARNESS FOR UP-TO 4 STACKED BEAMS*

YAGI STACK SEPARATION TEST

INTRODUCTION

THE STACKED ARRAY

MATCHING AND PHASING HARNESS FOR STACKS TOP-BOTTOM-STACK TESTS*

REMOTE, TOP-BOTTOM-STACK SWITCHING

STACKING TRI-BAND YAGIS

3X3 GK stacks rotating tower

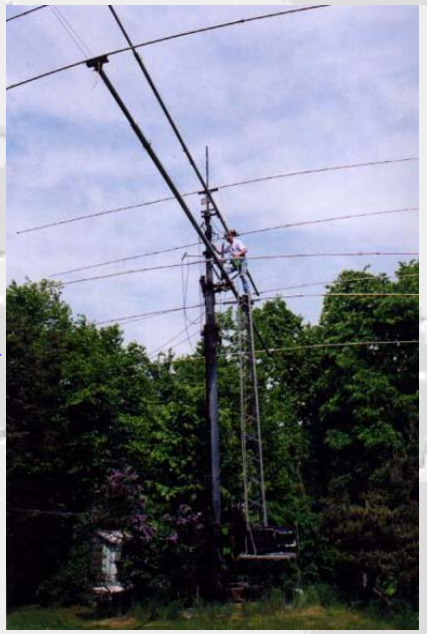
MODIFYING THE 204BA

6 EL 20M YAGI ANTENNA PARTY

THE 50 FT BOOM 5 EL 20M YAGI ON IT'S WAY UP TO 100 FEET

CONTEST HF AMP 2, 4CX1000A IN PARALLEL VACUUM EVERYTHING

6 EL 20M YAGI ANTENNA PARTY (note; the reflector is on the ground) 60 FT LONG 4 IN DIA BOOM



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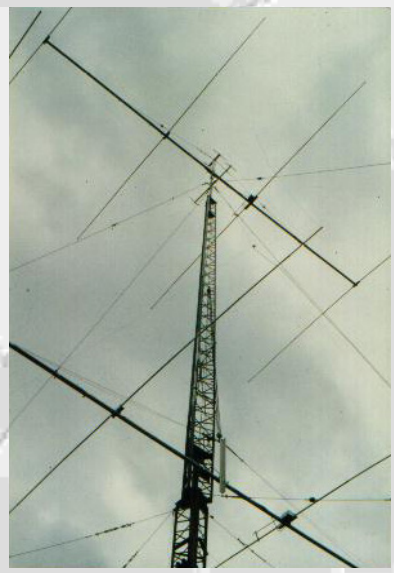

4X4 20M STACKS ON A MODIFIED LATTICE TOWER 90 FT AND 40 FT

THE BRONTOSAURUS 20METER GUN

4,5,6 and 7 element 20m VE3GK yagis

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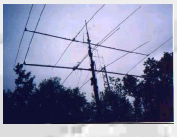
UP 100 FT

THE COMPLETED 4X4 20M STACKS ON THE 2.5 TON TELESCOPIC ROTATING SKYNEEDLE



DOWN 28 FT

THE RED 4CX1000A HF AMP

LAST UP-DATE, 07/23/02

HF TRANSCEIVERS

1.5 TON STEALTH TOWER

winch in the winter

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2.5 TON 115 FT SKYNEEDLE ROTATOR AND 2 HP HOIST

THE 50 FT BOOM 5 EL 20M YAGI ON IT'S WAY UP TO 100 FEET

SPECIAL PROJECTS

204BA CONVERSION TO THE 204GK**

YAGIS

4 EL 20 M YAGI

6 EL 20 M YAGIt

5 EL 20 M YAGI

7 EL 20 M YAGI

4X4 20 M STACKS 1

4X4 20 STACKS 2

HOW ANTENNA TRAPS WORK

20 M STACKS AND ROTATING TOWERS

HB SKYNEEDLE TOWER 2.5 TONS

1.5 TON 80 FT SKYNEEDLE ROTATOR-TILT-OVER- HOIST

20 METER 4X4 STACKS 42 FOOT BOOMS ROTATING 90 FOOT LATTICE TOWER

BIG TOWER ROTATORS

ROTATING LATTICE TOWERS

HF AMPS

A,,,4CX1000A 6 METER

20M GK YAGIS

4 el,*.. 5 el,*.. 6 el,*.. 7 el *

204BA CONVERSION TO THE 204GK**

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SPECIAL PROJECTS

STACKING TRI-BAND YAGIS

THE GK HOCKEY PUCK NOISE SUPPRESSOR FOR YAGIS

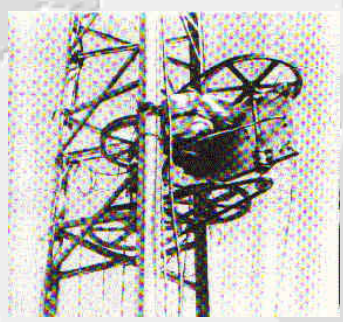
20 METER 4X4 STACKS*

YAGI STACK SEPARATION TEST

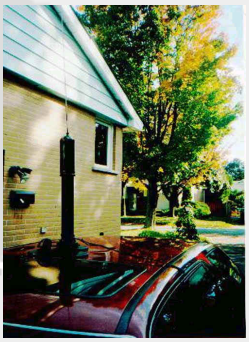
REMOTE, TOP-BOTTOM-STACK SWITCHING

GAIN GAME STUFF*

LINEAR TO CIRCULAR ROTATOR



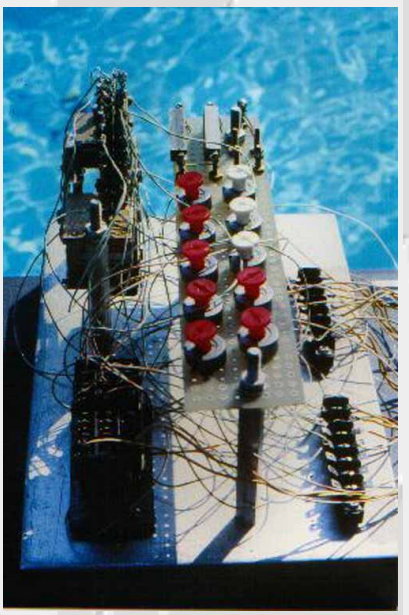
VEE BELT ROTATOR



MAG MOUNTED MOTORIZED 3 TO 30 MHZ < 1.1 TO 1 VSWR WITH NO EXTERNAL L/C COMPONENTS THIS IS NOT A COMPROMISE SCREWDRIVER ANTENNA

2.5 TON 115 FT SKYNEEDLE ROTATOR AND 2 HP HOIST

ve3gk@rogers.com



VHF-HF-UHF REMOTE CONTROL BEAM CONTROLLER

ve3gk@rac.ca

NB:

If you build anything, follow the technical info, if you change the info the results may not be positive.. GER

+

Please follow good safety rules when you climb towers and work with high voltage power supplies.

BUILDING THE TOWERS^^

B,,,4CX1000A HF.TABLE TOP

C,,,4CX1000A MINI SIZE

D,,,4CX1000A

E,,,4CX1000A

F,,,4CX1000A

G,,,4CX1000A A PAIR 1.5-30MHZ CONTEST AMP

PLI72 1.5 KW

PLI72 STATION

4X150D AMP AND HF TRANSCEIVER

4CX1000A AND HF TRANSCEIVER

HF TRANSCEIVER

7 HF AMPLIFIERS and 4 HF TRANSCEIVERS*

SPECIAL PROJECTS

MISC PROJECTS

GUESS WHO ONTHE BOOM

THE GK HOCKEY PUCK NOISE SUPPRESSOR FOR YAGIS

QUARTER-WAVE COAX MATCHING HARNESS FOR UP-TO 4 STACKED BEAMS*

MATCHING AND PHASING HARNESS FOR STACKS TOP-BOTTOM-STACK TESTS*

don't skimp on the stack separation

GK HF STATIONS

A LOOK AT THE DX LOCATION CLICK HERE

Guess who on the boom

VE3GK 20 DB, HF, 160 TO 10 METER AMP USING TWO 4CX1000A CERAMIC TUBES

Building the big tower in the shop

TUNED TWO METER CAVITY, BAND PASS FILTER TO REDUCE INTER-MOD

3-30 MHZ MOTORIZED MOBILE ANT

PL172 HIGH RANGE CONTEST CONTINUOUS DUTY 3 TO 30 MHZ AMPLIFIER

GK HF TRANSCEIVER + 1 KW AMPLIFIER miniature

THE 50 FT BOOM 5 EL 20M YAGI ON IT'S WAY UP TO 100 FEET

THE 5 ELEMENT YAGI CAPER

BIG TOWER OUT FOR MAINTENANCE... SOCKET IN THE BACKGROUND

4EL 20M GK YAGI 40FT BOOM

My son Michael, above, working on the 2.5 ton big tower, nested, 118ft when extended

GK HF STATION 3-30 MHZ TRANSCEIVER + 4CX1000A AMPLIFIER

3X3 GK stacks rotating tower

CHANGING A SLAVE HOIST CABLE

LINEAR TO CIRCULAR HEAVY DUTY ROTATOR

The stacked array ant direction controller

BIG TOWER REMOTE SAFETY CONTROL

RELAY SYSTEM

HOMEBREW GK HF STATIONS

THE BIG TOWER NESTED ST 28 FT...MY SON MICHAEL WORKING FROM THE WORK TOWER, CHANGING ONE OF THE OMEGA MATCHES

GUEST SPOT#2

F6DMQ.... STACKING TWO KT34XA YAGIS ON A UNIQUE ROTATING TELESCOPING ROTATING TOWER

PICTURE OF HALF DEPLOYED HDX 589 TOWER

PLEASE SEND ME IMAGES OF YOUR HOMEBREW STUFF 73 GER VE3GK

GUYS FOR GUYS AND GALS WHO HAVE TO GUY

TIPS ON HOW TO EXTEND THE LIFE OF TRANSMITTING TUBES

HOW ANTENNA TRAPS WORK

GK HF STATIONS

A LOOK AT THE DX LOCATION CLICK HERE

CHANGING A SLAVE HOIST CABLE

The stacked array ant direction controller BIG TOWER

Building the big tower in the shop

TUNED TWO METER CAVITY, BAND PASS FILTER TO REDUCE INTER-MOD

My son Michael, above, working on the 2.5 ton big tower, nested, 118ft when extended

LINEAR TO CIRCULAR HEAVY DUTY ROTATOR

BIG TOWER OUT FOR MAINTENANCE... SOCKET IN THE BACKGROUND

MOTORIZED MAG MOUNT MOBILE ANT. NO INTERMEDIATE LC NETWORKS <1.2/1, 3-30MHZ *

GAIN GAME STUFF*

BIG TOWER ROTATORs

MOTORIZED MAG MOUNT MOBILE ANT. NO INTERMEDIATE LC NETWORKS <1.2/1, 3-30MHZ *

6 EL 20M YAGI 58 FT BOOM you think the antenna is big you should see the size of the TV set

4X4 20m stacks

3X3 20m stacks

4X4 20 M STACKS HB SKYNEEDLE TOWER

2.5 TONS

back to the top

BUILDING THE 1.5 TON TOWERS ^

Z

TOWER REMOTE SAFETY CONTROL RELAY SYSTEM

back to the top of the frame

HF MOBILE ANTENNA (HB9ABX)

Felix Meyer

update 27th May 2003 / details July 27, 2004

This mobile antenna is designed for all HF bands from 10 to 80 meters an proved to be very efficient in my travels in South America on 15 and 20m for contacts to Europe and within South America.
The antenna is my own development.
Compared to commercial HAM Antennas the performance was always much better, due to the larger size and lower loss in coil.

Comparing with the Screwdriver type antennas always resulted in 2-4 S-points advantage in favor of this antenna!

In field tests we compared 3 different mobile antennas, all at the same location, at the same time, each one with 100 W power.

The 3 antennas were:

- HUSTLER mobile antenna (10 - 80 m)
- YAESU ATAS-100 (10 - 40 m)
- HB9ABX mobile antenna (10 - 80 m)

In all tests the signal of the home made antenna was the strongest!

- 1 S points (up to 10 db) stronger than HUSTLER
- 2 to 4 S points stronger than ATAS-100/ATAS-120

These results in the ground wave (5 to 10 km), as well as in the far field at 70 km to 1000 km distance.

For 10, 15, and 20 m the antenna consists of a fixed lower part and an extensible whip (telescopic antenna of 15 to 80 cm length) on the top.
For 20, 40 and 80m a second segment is added and at the top follows the whip.

Both parts are made of fiberglass rod of 10 mm diameter and 175/165 cm length, on which a piece of enameld copper wire (CuL) of 1.5mm diameter is wound. Aluminium tubes are used to join the parts together.
On both fiberglass rods equally spaced windings of emameld copper wire of 1.5 mm dia are wound. The lower part holds 320 cm wire (79 turns) and the upper part holds 470 cm wire (120 turns).

The loading coil is wound on a plastic tube of 7.5cm diameter and 17 cm length.
The same type of wire (1.5 mm CuL) is used for the coil as for the antenna.
The coil has 33 turns in total, coil length is 12 cm.
Seen from the bottum of the coil, there are taps at 21, 27, and 33 turns,

which are shorted according to the operating frequency.

Settings of the antenna in operation:
10m : Lower part + whip + tap 3 + bridge from tap 3 to the whip
15m : Lower part + whip + tap 3
20m : Lower part + whip + tap 2
20m : Both parts + whip + tap 3
40m : Both parts + whip + tap 1
80m : Both parts + whip + no tap (full coil)
The WARC bands 30, 17, and 12 m can be operated by just adding coil taps without changing the coils.

Feeding:

The antenna is fed by 50 Ohm coax (Type RG 58) of about 2 m length.
On the antenna side the center conductor of the coax is connected to the antenna wire and the coax braid is connected to the car chassis ground.

The antenna system requires a good ground !
I soldered a flexible wire at both sides inside the door frame leading to the coax braid.

Don't forget to connect the two windings on the fiberglass rod together, either through the metal sleeve between the rods, or simply by a connector. The whip on the top is connected to the end of the wire below.

The desired resonance frequency is adjusted by changing the length of the whip on top of the antenna. Extending the whip lowers the frequency.

The antenna requires an antenna tuner for operation, as the impedance differs from 50 Ohms on 40 and 80 m.
With the aid of the tuner, the antenna is adjusted to SWR 1.0 on all bands.

The MFJ-901B tuner is fine for this use. It has little weight and is very small (12.5 x 5 x 15 cm).
(Before its use it's recommended to open it to check the adjustment of the variable capacitors. They are frequently bad centered, which produces easily shorts ...)
If the transceiver and tuner are without SWR meter then a separate instrument is to be connected between TX and tuner. I am using a DAIWA CN410M which is a small cross needle instrument best suited for this use.

Mounting of the antenna:

**I installed the antenna on top of the car on a roof carrier.
Two aluminum angles, 30 cm length, 30mm thigh, were fixed
with screws to the carrier and the rod below the coil fixed to the angles
by means of two hose clamps.**

**Instead of the Aluminium tubes copper tubes may be used, whatever
is found to be better. The tubes are glued in 5 cm length over the
fiberglass rod. At the end of the 7 cm open part of the tube, a 2 cm
long slot is sawed and a small clamp is used to hold the inserted antenna
part .**

**The antenna is directed backwards at an angle of abt. 70 degrees.
In order to prevent swinging up, the antenna is fixed slightly down
at the end with a nylon rope. Use 2 nylon ropes, to both sides to prevent
swinging sidewise when driving. This is required when both antenna
elements are installed.**

**The top of the antenne nearly reaches 3.2 m height above ground when
both elements are in use and fixed by a nylon.**

Adjustment hints:

**Initially, the antenne has te be tuned to resonance on each band
without using the tuner.
This adjustment is done by changing the coil tabs and then varying
the spacing between turns.
In operating mode, only the tab setting is changed, the whip length set,
and tuned with the tuner.**

Before initial adjustment, both elements have to be checked: click here

**The initial adjustment is done by using an antenna analyzer (e.g.MFJ-259).
Resonance is found by obtaining minimum SWR on the instrument.**

Procedure of initial adjustment : (this is done without tuner)

Lower part + whip (10 to 20 m)
- Coil tap 3 (= all turns closed)
Adjust length of whip to obtain min. SWR at 21.2 MHz.
Note length of whip. This is the 15 m setting.

- Coil tap 2 (27 turns closed)

Keep length of whip from previous setting.
Adjust spacing of the upper 6 turns to obtain min. SWR at 14.2 MHz.
Fix coil, this is the 20 m setting.

- Coil tap 3
Make a bridge using a flexible wire of 150 cm length with clips between tap 3 to whip connection (wound 2 - 3 times around rod).
Adjust length of whip to reach min. SWR at 28.4 MHz.
Note length of whip. This is the 10 m setting.

Both parts + whip (20 to 80 m)

- Coil tap 3
Adjust length of whip to reach min. SWR at 14.2 MHz.
Note length of whip. This is the 20 m setting.

- Coil tap 1
Measure SWR on 7.05 MHz and adjust spacing of coil between taps 1 and 2 to reach minimum SWR. Change length of whip only, if minimum SWR remains outside indicated frequency.
Fix coil and note length of whip. This is the 40 m setting.

- Coil without tap (33 turns active).
Keep length of whip and measure SWR on 3.7 MHz.
Adjust spacing of coil of the first 21 windings to reach miminum SWR.
Change length of whip only, if min. SWR remains outside 3.7 MHz.
Fix coil and note length of whip. This is the 80 m setting.

After terminating this adjustment the coil winding is fixed permanently.
After this initial adjustment, the tuner is connected to mach the SWR exactly on operating frequency.

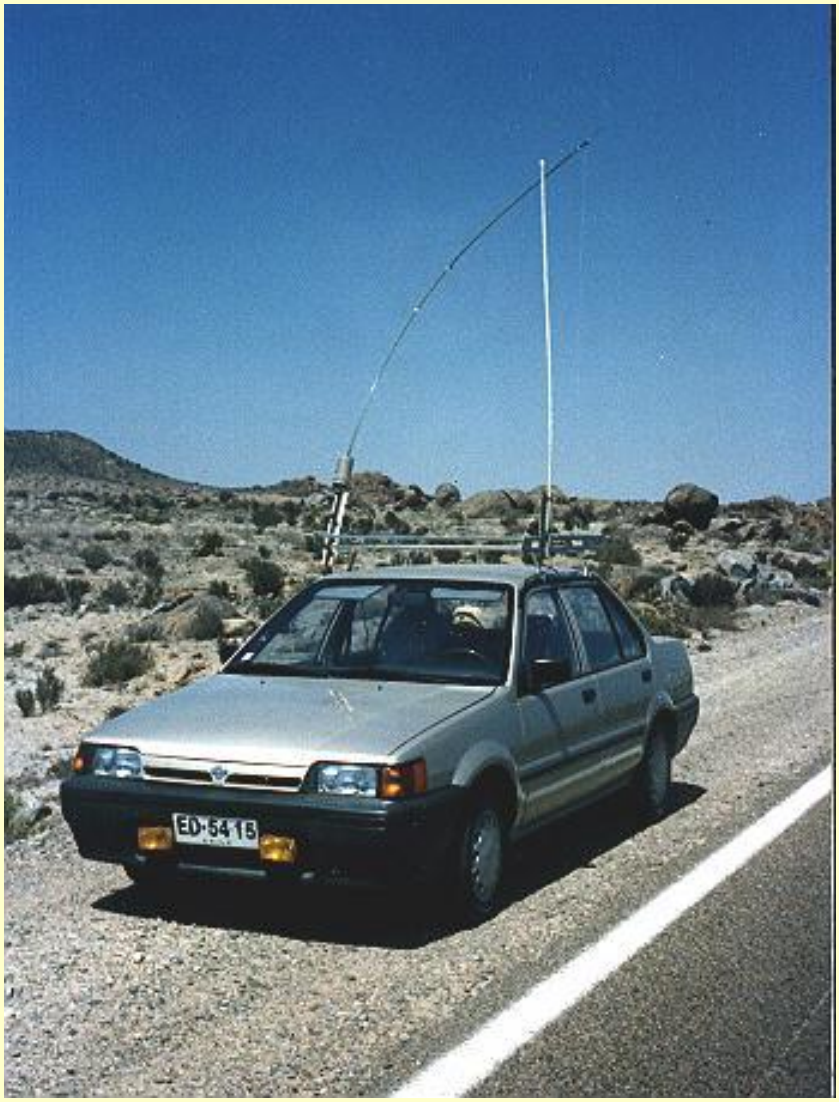
With my antenna the SWR without tuner on 10, 15, and 20 m is better than 1.2 therefore tuning is only required for 40 m and 80 m operation. This however depends from the actual installation, especially from the ground connection.

See further details from the following drawings:
Drawing 1 (lower part)
Drawing 2 (upper part)

The photograph below shows the installed antenna in Chile as CE3CWF/mobile.
(The white antenna is a 2m antenna)

Good luck in construction.

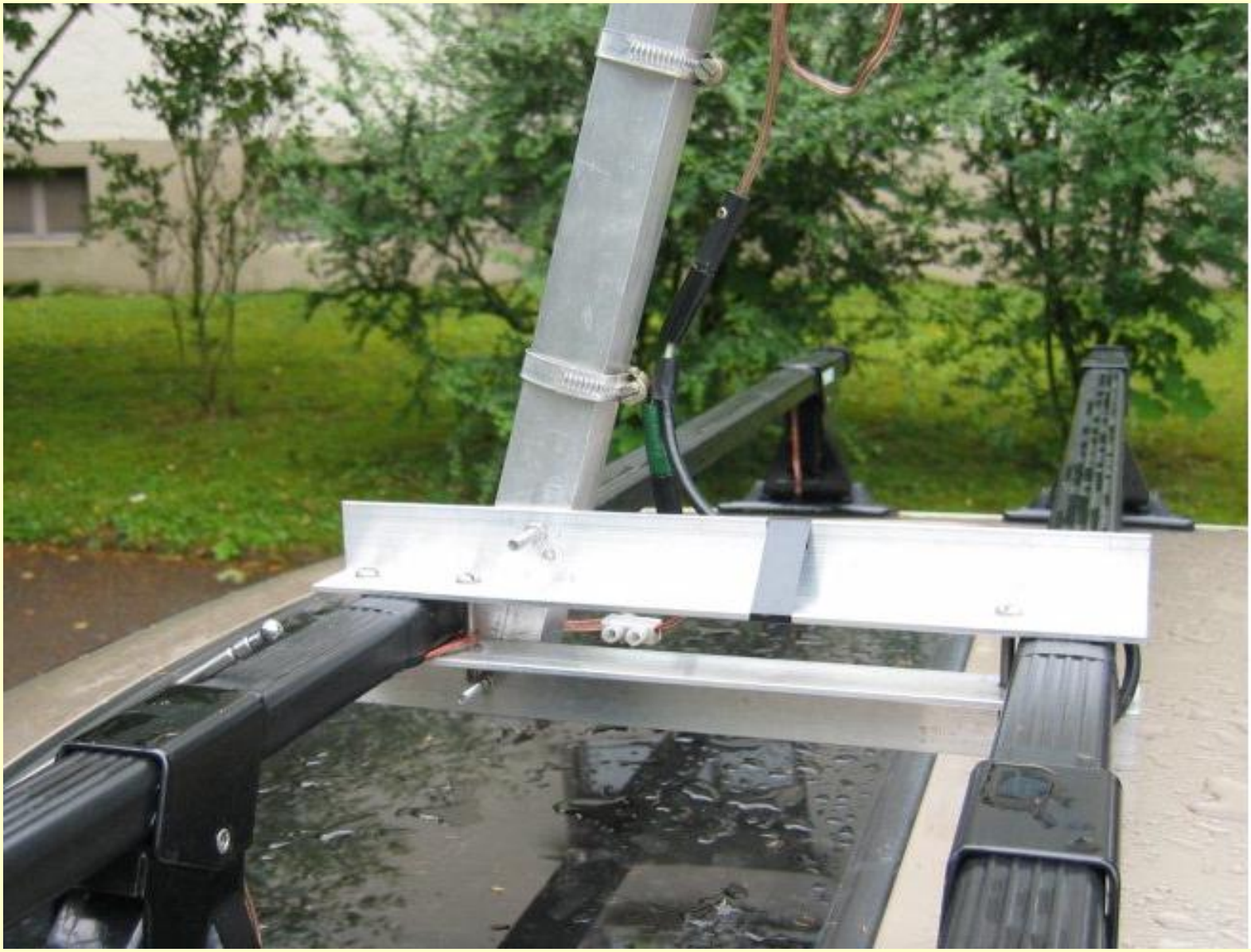
For additional questions: see here



my antenna as CE3CWF

Below the mobile antenna in Switzerland as HB9ABX:

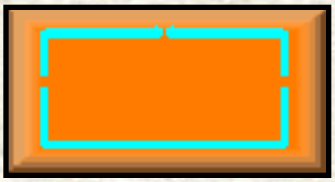
And finally the mounting of the antenna on the roof of the car:



WARNING

Working with fiberglass rods produces dangerous dust, which easily produces allergies on the hands. Inhalation of the dust is extremly dangerous !
Already holding the raw bars with the hand can produce problems, therefore use protecting gloves
whenever working and handling raw fiberglass rods and observe appropriate care when sawing.
Wash well the hands with soap after the work !
After terminating the work paint the complete rods with a suitable lacquer to protect the surface.

BACK



Moxon Rectangles

L. B. Cebik, W4RNL

The Moxon Rectangle is growing in popularity as a compact 2-element array that approaches a full-size 2-element in gain but with a far superior front-to-back ratio and a direct match for the standard 50-Ohm coaxial cable. The antenna can be built as a wire array--especialy for the lower HF regions--or as a rotateable alumunum beam. For convenience, I have pulled together the growing selection of Moxon Rectangle notes into this single subdirectory and organized them in the order of recommended reading--unless you already know what you are looking for.

- **VK2ABQ Squares and Moxon Rectangles** (7-25-99)
- **The Moxon Rectangle: A Review** (12-18-99)
- **An Aluminum Moxon Rectangle for 10-Meters** (4-6-98)
- **Wire Moxon Rectangles for 40-10 Meters** (1-1-97, 6-10-99)
- **Further Notes on 40-Meter Wire Moxon Rectangles** (5-4-99)
- **Notes on the Moxon Rectangle Pattern** (2-22-2000)
- **The Double-D Antenna** (3-18-97, 5-4-99)
- **Multi-Banding the Moxon Rectangle** (8-12-99)
- **The Moxon Rectangle on 2 Meters** (10-01-99)
- **Building a 2-Meter Moxon** (12-18-99)
- **Designing Moxon Rectangles by Equation and by Model** (10-01-2000)
- **40 + 30 = 50 (Not 70)** (10-23-2001)

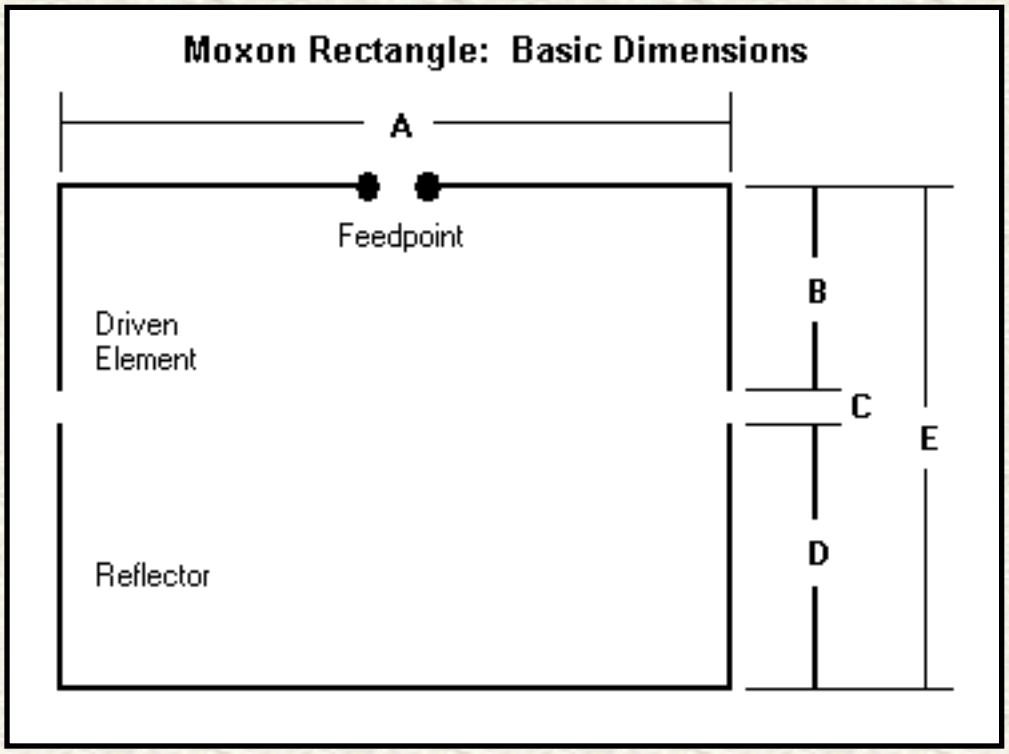
- **A Truly Portable Moxon Rectangle for Nearly No-Tool Field Assembly** (02-01-2003)
- **Moxon Rectangles for 6 Meters** (02-03-2003)
- **The Elusive Moxon Nest** (06-01-2003)
- **Part 1: Vertically Stacking Horizontally Oriented Rectangles** (12-04-2003)
- **Part 2: Vertically Stacking Vertically Oriented Rectangles** (12-04-2003)
- **Understanding the N0KHQ Coax Square** (12-08-2003)

Additonal information on building wire and tubing versions of Moxon rectangles for a direct 50-Ohm feed is available in ***Simple and Fun Antennas for Hams***, ed. Hutchinson and Straw (ARRL, 2002), pp. 12-19 to 12-28. The **KD6WD Moxon Antenna Project** at Murray State University is another good source of information on various construction techniques, especially for the operator needing a light-weight or a semi-stealthy antenna.

MOXGEN: A stand-alone Windows program that calculates the dimensions of a Moxon rectangle for a near-50-Ohm feedpoint impedance, as described in "Designing Moxon Rectangles by Equation and by Model," has been developed by Dan Maguire, AC6LA. You may obtain a free copy from his website: http://www.qsl.net/ac6la/moxgen.html. The program will also create a model in .EZ format for use with EZNEC or in .NEC format for use with NEC-Win software or with generic NEC programs. The only required input entries are the design frequency and the diameter of the wire or tubing to be used. Dan's site also contains a number of other very useful programs for modelers and others interested in antenna design and analysis.

Below is a version of the Moxon dimension calculator that you may use right on this page, thanks to Joe Faber, KG4UHP, who created the JAVAScript and gave me permission to place it here. Remember that the dimensions apply to Moxon rectangles that use the same diameter material throughout. Decide on the design frequency and the diameter of the elements. You may use inches or millimeters for the diameter--or you may select an AWG wire gauge. Be certain to select the unit of measure for the output. Then, click on any of the output boxes if the calculations have not already appeared.

Moxon Rectangle Dimension Calculator



Frequency : MHz

Wire Diam :

Output Units :

A

B

C

D

E

Updated 12-08-2003. © L. B. Cebik, W4RNL. Data may be used for personal purposes, but may not be reproduced for publication in print or any other medium without permission of the author.

Return to Amateur Radio Page

Background
Participants
Design
Construction
Performance
Msg Board
Links

The aim of this website is to share the experiences of MOXON antenna builders everywhere. It's all about Hams Helping Hams which is the overlying theme of the MOXON Antenna Project.

The MOXON is a stealthy gain antenna that can be easily home brewed for a small investment. This antenna can be mounted on a simple push-up mast and rotated by hand or with an inexpensive light duty rotor. It should be especially attractive to hams who want to avoid purchasing more expensive towers, rotors and directional antennas.

We invite you to click on the various links at left to learn more about the project.

73,

Moxon Antenna Project Participants

p.s. Want to include your MOXON on this website? Just email your construction notes, performance observations and photo attachments to the website author *Don Snodgrass*, K4QKY.

Click HERE to post an email to the MOXON message board

6 meter MOXON inquiry and response added to the MOXON Message Board. Click here.

W3CRR 20 meter "yagi style" aluminum tubing MOXON. Click here.

9A2TD 20 meter MOXON. Click here.

New posts on the MOXON Message Board. Click here.

N2YET modified 17 meter MOXON. Click here.

9Y4DD 20 meter MOXON. Click here.

DF4RD provides preliminary info on his 12/17 meter "HEXMOX" Click here.

Inexpensive source for Crappie fishing poles (spreaders) found. Click here.

MOXON message board added to website. Click here.

F5MAG multi-band hybrid MOXON. Click here.

N5GLR 20 meter hybrid Coaxial MOXON. Click here.

YU1QT 6 band hybrid MOXON. *(recently updated)* Click here. YU1QT has also built a 3 element 40 meter MOXON which you learn about by clicking here.

W1ZY 40 meter MOXON. Click here.

KA4SDU 17 meter Bamboo MOXON. Click here.

K06HL 3 element Hybrid MOXON. Click here.

Free Web Counters

Kohls Store

MOXON Antenna Project web pages hosted courtesy of
Murray State University Amateur Radio Club

Site Author: Don Snodgrass, K4QKY
http://webpages.charter.net/donsno

Petlowany Three-Band Burner Antenna

June 23, 2002

The *Petlowany Three-Band Burner* is a simple, low-cost, trapless short vertical antenna which amazingly works on three HF bands (20, 15 and 10 meters). This web page contains pictures, performance data, and enough construction details so you can "homebrew" your own.

Petlo... Who?

A ham named Bill Petlowany (K6NO) published an article in the March 1998 issue of Worldradio magazine, reporting some interesting results from placing spiral coils on the ends of antennas. A handful of others have experimented with these peculiar antennas. (Another article by W. Caldwell (WA8ABE) appeared in the June 1999 issue of Wires and Pliers.) I kept *Petlowany* in the name of my variant in honor of the original experimenter, and called it a *burner* because of its shape.

Basic Description

Fundamentally it's a quarter wave ground plane with 4 radials cut for 15 meters. The interesting twist is the spiral coil "hat" on top, which makes the antenna resonant on 20 and 10 meters as well. The higher wires you see in the picture are guy wires, broken into small enough pieces that the RF doesn't see them. They're not really part of the antenna, electrically speaking.

Performance-wise, it is pretty much indistinguishable from a quarter wave ground plane vertical on all three bands. As such, it does a reasonable job of delivering the low angle radiation required for DX operation.

SWR Measurements

Here are some SWR charts, to demonstrate the antenna's performance on various bands. These are actual measurements of the antenna pictured.

Here are the three bands that give the *Petlowany Three Band Burner* its name.

- SWR chart for 20 meters (1.4:1 at 14.0, 1.2:1 at 14.5)
- SWR chart for 15 meters (1:1 at 21.0, 1.1:1 at 21.5)
- SWR chart for 10 meters (1.2:1 at 28.0, 2:1 at 29.3)

Here are two more bands that will at least accept power. Performance isn't great on these bands, but it's usable in a pinch.

- SWR chart for 80 meters (2:1 at 3.5, 2.5:1 at 4.0) SWR is high but it won't cook your finals. For an antenna this short to do *anything* on 80 is pretty nifty.
- SWR chart for 2 meters (1.1:1 at 144, 1.5:1 at 146) Don't let the low SWR fool you. The vertical is so long for 2 meters that most of the energy goes up at a 60 degree angle.

Construction Details

Following are detailed instructions for building the antenna as I have built it. Naturally many other types of construction are possible, and different dimensions could yield antennas that work on other frequencies.

Parts List

Most everything can be found at a hardware store and a radio shack.

- 12 feet of 1/2 inch PVC pipe (thicker variety preferred)
- one 1/2 inch PVC cross
- one 3/4 inch PVC male plug
- 5 meters of 12 ga. wire (recommended: stranded, insulated)
- 20 cm of 3/4 inch copper pipe (as above, the thicker kind is apt to be stronger)
- two 3/4 inch copper male adapters

- two 3/4 inch copper pipe mounting brackets
- one 10 foot length of one inch copper pipe (In hardware stores, two grades are available. Get the thicker one, "blue" or "type L")
- one 1 inch copper male adapter
- one 3/4 inch thread to 1 inch copper male adapter
- 2 radio shack guy rings
- a few feet of solid wire (14 ga or 12 ga, not critical)
- one 5 foot galvanized pipe (one inch) with threaded ends
- two inches of one inch PVC pipe (thicker variety preferred)
- two 1 inch PVC female adapters
- 20 meters of 12 ga. wire (I used stranded, bare wire from radio shack, but the exact type is not crucial)
- 100 feet of steel guy wire (available at radio shack)
- 1 radio shack roof apex antenna mount kit
- lag screws suitable for attaching roof mount to roof
- roof tar to seal over roof mount
- 4 large hooks for guy wires (available at radio shack or hardware store)
- coax

Making the spiral top

Click the picture above for a more detailed view.

The spiral coil is wound on a form built of half inch PVC pipe. The first step is to make the coil form. Cut four pieces of half inch PVC 20 cm long. Insert one of them into the half inch PVC cross as far as it will go. Make a mark on the pipe where you would be able to drill a 3/16 inch hole as close as possible to the cross without touching it. This is the hole through which the first turn of the spiral coil will pass.

Remove the pipe from the cross and make additional marks every 2 cm all the way out to the end (the long way, away from where the cross fits). Use the first pipe as a template and mark the other three pipes the same way. Now drill 3/16 inch holes as marked, all the way through both sides of all four pipes.

Using PVC cement, glue the four pipes into the cross to make the coil form. Be sure to glue the right ends of the pipes, so the innermost hole is close to the cross without being obscured by it. Take care also to keep the holes in line with the plane of the cross, so a coil can be wound, as in the picture.

Use 5 meters of 12 ga. wire for the spiral coil. Any type of wire is suitable electrically. I used wire with a slick insulation for ease of threading through the coil form. Starting from the inside, leave 10 cm of slack for the inner coil connection, and thread the wire through all four innermost holes in the coil form, pulling it tight.

Continue to thread the wire progressively outwards in a spiral shape, pulling it tight as you go. Make sure you always have the 10 cm of slack inside for the connection to the pipe. When done, use a cable tie to keep the wire from slipping back out, as shown. To reduce wind resistance, you may wish to saw off any excess length on the coil form arms. (The ones in the picture have not been trimmed.)

The final step in making the spiral coil is attaching the 3/4 inch male PVC plug to the side of the cross, so the coil can be screwed down onto the copper pipe. To do this, rough up one side of the cross with sandpaper. If you will do most of your communication in the Northern Hemisphere, let the coil run clockwise... (just kidding!). Also rough up the flat end of the 3/4 inch PVC male plug. Slather on some PVC cement and hold the plug in place until the glue has a chance to set up. Here is a close-up picture of the joint between the plug and the cross. It may not look strong but PVC cement forms a strong chemical bond. Do not screw the plug into anything until the glue dries overnight.

Now that you have the coil, you must make the pipe portion of the antenna top. This is easy. Sweat a 3/4 inch female adapter onto each end of a 20cm length of 3/4 inch copper pipe. (If you wish to experiment with tuning, you might want to leave one end unsweated (as shown) and/or start with a longer length of pipe. If you wish only to duplicate my results, don't worry about getting the length perfect -- the tuning is very forgiving, in terms of the length of the vertical element.)

As you can see in this detailed photo, I sweated on mounting brackets near the top end to make available four holes at right angles. Once I'm sure I won't want to take down the antenna for a while, I might use those to attach ropes to steady the top of the antenna. For now, though, I prefer not to use the ropes. It is so much easier to take the antenna down without them. Even if you don't plan to use ropes, you must sweat on something that will allow you to solder the spiral coil wire without damaging the PVC plug on the coil form with excessive heat.

To complete the spiral top assembly, screw the coil form into the pipe, with the mounting brackets (or other soldering post) toward the top. Cut off any excess length and solder the inner end of the spiral coil to the bracket or post. Be careful not to damage the PVC plug with heat.

When completed, your spiral top should look a lot like the one in the photo.

Making the main vertical element

Click the picture above for a more detailed view.

The vertical element is chiefly a 10 foot length of one inch copper pipe. The only critical thing about this assembly is that it is necessary to slide a guy ring over one end of the pipe *before* sweating the top end

on.

Choose an end of the pipe to be the bottom and sweat on a one inch male adapter. Also sweat a piece of wire (I used 14 ga. solid but it is not critital) to the bottom end for attachment to the feed line. In this detailed photo, the wire can be seen as a ring just above the male adapter sweat joint, and also where it attaches to the coax.

About six feet from the bottom end, sweat on a ring consisting of a few turns of solid wire. This ring will stop the guy ring from sliding down. (The picture at the beginning of this section shows this ring stopper in detail.)

Slide a guy ring onto the top of the pipe, making sure it is the right way up (skirts point down). Be sure the ring you just soldered onto the pipe holds the guy ring in place. Also make sure the guy ring can turn freely while straining against the stop. This will make it much easier to install and take down the antenna.

Finally, sweat a 3/4 inch male adapter (for 1 inch pipe) to the top end of the pipe. This will attach to the spiral top when the antenna is assembled.

Preparing the base

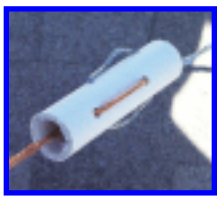
Click the picture above for a more detailed view.

The base of the antenna consists of a galvanized pipe which acts as a mast and a PVC "barrel" which secures the antenna to the mast mechanically without connecting it electrically.

To make the PVC barrel, simply cement two one inch PVC female adapters to a short length of one inch PVC. The pipe should be just long enough that the two adapters touch or nearly touch.

Before screwing the barrel onto the mast, it will be necessary to slide a guy ring on. If you use the recommended radio shack guy ring and one inch galvanized pipe, the guy ring won't *quite* fit over the end. Use a grinding or reaming tool (I used a dremel) to increase the size of the hole in the guy ring. It is not necessary to make it fit over the entire pipe. It only needs to be able to slide down over the threads enough that the barrel can screw on securely.

Slide the guy ring over the top of the mast, using the included bracket if necessary to keep it from sliding down. Screw the PVC barrel onto the top.



Making the radials and guys

Click the picture above for a more detailed view.

Both the radials and guys require insulators. If you have 24 good quality egg insulators, by all means use them. I had trouble finding any and decided to use PVC to make low-cost insulators. As you can see from the picture, I drilled four 3/16 inch holes through a 4-inch piece of half inch PVC, with interlaced pairs of holes at right angles. (No way that language would be clear without a picture!) The advantage to this approach is that the wires are actually linked, so when the PVC inevitably rots from the sun's UV rays, the structure is still mechanically secure. That's the theory, anyway. Note the way each wire is threaded into one end, out and over the side, then through to the other side, over and back in, then out the end again.

Once you have made or otherwise obtained 24 insulators, you are ready to prepare the radials and guys. We'll start with the radials.

To make the radials, cut four lengths of 12 ga. copper wire slightly longer than 3.5 meters, such that the radial will be 3.5 meters long once the ends are wrapped around the guy ring at the base and the insulator at the other end. Attach an insulator to one end of each radial. Solder the wire as shown for extra strength.

Attach the non-insulator end of each radial to the guy ring at the top of the mast (just below the PVC barrel). Do this by making a loop and soldering it as shown in this detailed photo. Solder a loose ring of copper wire to each loop, all the way around the guy ring, to ensure that the radials are electrically connected at all times. Coil up and tie the radials neatly. They are now part of the base, ready to install on the roof.

To make each guy, use steel guy wire to construct a string of five insulators spaced one meter apart, for a total length of 4 meters per guy string. (Longer, electrically continuous guy segments can mess up the antenna's performance.) Make 4 guys, using a total of 20 insulators.

Use the shortest length of steel guy wire possible to attach the insulator on one end of each guy string to the guy ring that slides freely on the main vertical element. This detailed photo shows how. Neatly coil up and tie the guy strings. These are now part of the vertical element, ready to install.

Installing the base

Click the picture above for a more detailed view.

Find a suitable location at the apex of your roof, surveying where the radials and guy wires will go. Avoid power lines for safety. Once you find a good spot, screw the roof mount down securely with lag screws. Use roof tar to weatherproof the mount as shown before the next rain or snow.

Install hooks for the guy wires and radials at four locations such that the radials will be spaced as evenly as possible (90 degrees apart), and such that the radials can be fully extended (3.5 meters from top of base) and not yet reach the hooks.

Place the mast near the mounting point. If an assistant is available, the mast can be held in mounting position. Run the radials out toward the hooks (they shouldn't reach yet) and secure them to the hooks with a length of steel guy wire extending from the insulator to the hook. This picture shows what the insulator looks like with one copper wire and one steel wire. Leave enough slack that the radials can be under tension, fully extended, once the mast is raised. Without an assistant, getting these lengths close enough to let go is cumbersome. Be patient and don't risk letting the mast get out of your control.

Once the mast is basically in place, adjust the tension at the hooks until the radials are fairly taut and the mast is vertical and feels secure. Tighten the mounting screws and lock nuts in the apex mount to complete the installation of the base.

Adjusting the guys

When the guys are adjusted to the perfect length, one person over five feet tall can easily raise and lower the antenna without help. It is necessary only to push the antenna up into the guys, then screw it down into the PVC barrel. Until the guys are perfectly adjusted, though, it can be dangerous to attempt to raise the antenna fully without assistance. Proceed cautiously and have patience, and make all guy length adjustments without the spiral top installed.

Attach fairly long pieces of steel guy wire to the end of each guy string, and secure them to the hooks. It

should be possible to raise the vertical element near to its mounting position without straining the guys. Now slowly and patiently, progressively shorten the guys at the hooks until they are just long enough to permit the vertical element to be raised into position and screwed into the base. (Having an assistant is extremely helpful.)

Completing the installation

Now that the guys are adjusted, raising the antenna is very easy. Screw the spiral top onto the vertical element, then hoist the antenna into place and screw it into the base. Connect coax as shown in this picture, without soldering yet, and run it to an analyzer, noise bridge, or transmitter with SWR meter.

The only critical adjustment is the length of the coil, which mainly affects performance on the 20 meter band. 5 meters is a bit too long for the spiral coil wire, so you should find that the antenna is initially resonant somewhat below 14.2 MHz. Carefully lower the antenna repeatedly, shortening the spiral coil to bring the antenna into resonance near 14.2 MHz. I ended up with a 4.75m coil length. If you were able to duplicate my design parameters, you should find the performance on the 15m and 10m band to be as shown in the SWR charts.

About N5IZU

I got my first ham ticket in 1978 at the age of 12, and I've been building the occasional antenna ever since. Most of them haven't worked, but every once in a while, something interesting happens. When I read about Petlowany's spirals I just had to know more about them. I had no idea what would actually be possible.

I still feel I know so little. I've put this project on the web in the hopes that someone will write me a breathless email one day, explaining, "No, -DT, you've got it all wrong! A little change here and there and you can have *five* bands!" Even more exciting would be to learn exactly how and why it works the way it does.

In the meantime, it is nice to just enjoy it for a while. Recently I've been content to just haunt the bands, chasing whatever DX gets snared in the *burner*.

Go

to -DeeT's Hacks Page.

David B. Thomas (dt@dt.prohosting.com)

Pfeiffer Quads

KQ6RH

(C) 1998, 1999, 2000

Ray Jurgens

(Up-Dated 3/07/2000)

Pfeiffer Quads:

The Pfeiffer Quad was described in QST and has the desirable feature that it is smaller than a standard cubic quad but does not compromise performance very much. The Pfeiffer design is rather difficult to construct because of the complicated linear loading networks required. However, these complicated structures are unnecessary and do not add much, if anything, to the performance. The Pfeiffer quad has linear loading stubs at each corner that are pulled into the center toward the hub forming a Maltese Cross. These could just as well be replaced with inductors or transmission line stubs. In fact, the losses on 450 Ohm ladder line are small to insignificant, so Pfeiffer's complicated stub assembly can be replaced with simple transmission line stubs with no loss of performance. The only complication is in the electromagnetic modeling; here the density of pulses must be very high to accurately handle the closely spaced wires. The stubs can be stretched along the spreaders and attached to the antenna wires at each of the tips. In this case, 8' spreaders are convenient, so all that is required is to find the proper length of the ¼ wave shorted stubs that act as inductors to bring the loop into resonance. These stubs can be made slightly longer for the reflector and shorter for the director if one is used. Placing the 450-Ohm ladder line along the spreader slightly lowers its impedance and lengthens its electrical length by the same proportion. The 8' spreaders can be used to cover 20, 17, and 15 meters. In the 15m case, the stubs are only a few inches long and for 20 meters they are about 3/4 of the length of the spreader. If terminal strips are mounted at the end of each tip, the stubs can be changed easily to accommodate various bands. Some experimentation may be required to get the length of the stubs set, but once set, assembly is very quick for this quad. Our **Products Page** lists the our design document which gives the a complete set of graphs and design data as well as full construction details for our modified Pfeiffer Quad.

The Pfeiffer Quad is normally operated in the X configuration. The impedance of this antenna will be lower than for a full size quad, and some matching network may be required. The

version presented in QST used a gamma match on one of the stubs to drive the loop. For example, at 20 m, the stubs must to be about 74" long, and the driving impedance of a single loop is about 47 Ohms. Thus, a single loop is a good match to 50-Ohm line. The addition of the parasitic elements further lowers the impedance, and depending on the spacing of the loops, the impedance could drop to 25 Ohms. For further details on the original Pfeiffer Quad, check the following reference:

Andy Pfeiffer, K1KLO, "The Pfeiffer Quad Antenna System," QST, March 1994, pp. 28-31.

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Phased Arrays of Short Vertical Antennas

It is possible to devise an array of two or more vertical antennas to provide useful gain and directivity over a single vertical element, but proper design and adjustment may be a more complicated matter than one might think because gain depends on such factors as spacing between elements, phase difference, etc., and these will change from one band to the next. About the most one can hope to do is to work out a compromise design that will be truly effective on two adjacent bands.

Probably the simplest system would involve two vertical antennas spaced from 1/2 to 3/4 wavelengths apart at the highest frequency of operation and fed with equal in-phase currents. Such an array would be bi-directional and provide up to four decibels of "broadside" gain. At half the design frequency the broadside gain would be less than two decibels. Thus, an array of the two elements for 40 meters (spacing between elements approximately 35 ft.) would provide a fair amount of gain on that band and perhaps a few decibels less on 80 meters. The same array could no doubt be operated on the higher-frequency bands, but the directivity pattern (if any) would change considerably, for the great spacing (in terms of wavelength) would cause the production of "end-fire" as well as broadside radiation.

In the case of close-spaced elements (1/8 to 1/4 wave) it is possible to produce a unidirectional end fire pattern by feeding the two elements with a phase difference of 90 degrees by means of an electrical 1/4 wavelength "delay" line. This arrangement produces a rather broad single lobe (cardioid pattern) in the direction of the element with the lagging current, and very good front-to-back radiation are possible with such an array. However, both the radiation resistance and the feedpoint impedance will be much lower than for a single vertical element, making the system more critical with respect to operating bandwidth and impedance matching, especially if each element is physically shorter than 1/4 wavelength at the operating frequency and if the spacing between elements is less than 1/4 wavelength. Close-spaced out-of-phase short elements should have the best possible radial system under each element so that the earth loss resistance doesn't account for the major part of the overall feedpoint impedance. If this precaution isn't observed the earth loss may wipe out all or most of the theoretical gain of the array.

Still another problem with close-spaced arrays using physically short elements is that the power-handling capability may be less than for a single element because the mutual impedance between the elements will raise overall circuit Q while reducing the radiation resistance of the array.

As with the relatively wide-spaced in-phase arrays, the close-spaced out-of-phase types may be operated on other bands where the spacing and phase difference will not be optimum and where the directive pattern can be expected to change from band to band.

In general, the Butternut HF6V-X and HF2V will be under less of a disadvantage in phased arrays than conventional multi-trap designs that use a progressively smaller portion of the available radiator at 7 MHz and above, but the problems of devising an effective array on more than one band will remain the same. Wide-spaced in-phase arrays are generally better behaved than the close-spaced out-of-phase types and show higher values of radiation resistance. This makes the ground/radial situation somewhat less critical, but more real estate is required for effective broadside arrays, especially on 80 and 160 meters.

For further information on vertical arrays one should consult a recent edition of the ARRL Antenna Book.

Hawaii Ham Radio Information Pages

- **HF Operation in Hawaii**
- **Topband Operation from KH6**
- **A Polar Projection Map from Hawaii**
- **HI-QRP, QRP in Hawaii**
- **QCWA Hawaii Chapter *(QCWA-194)***
- **Soil Conductivity in Hawaii**
- **VHF Simplex Operation in Hawaii**
- **WWVH in Hawaii**
- **Important Ham-Related Designations for Hawaii and the Pacific:**
  - **Grid Squares in Hawaii**
  - **Counties in Hawaii**
  - **CQ Numbers and DXCC Designations**
  - **IOTA Numbers in the Pacific**

Antenna Information

Rabbit Ears have been used to solve the limited space antenna problem of how to operate HF from a condo in Urban Honolulu. After much testing, this design has been working:

The Upright V Dipole for Limited Space

Near Vertical Incidence Antennas in Hawaii have been the source of much speculation. The NVI mode is ideal for HF communications statewide, critical to emergency communications where VHF simplex cannot span the distances required. Here is a modeling study of optimized NVI dipoles:

The 40 Meter NVI Dipole

A comparison of **Portable Antenna types** using EZnec modeling. Including dipoles, optimized 1/4 wave vertical systems, fan dipoles and inverted V's:

Common Portable Antennas Compared

The **"Stake Stick"** is a multi-band antenna for the higher bands designed for portable use such as backpacking. It uses full size, tunable elements on 20, 17,15 and 10 meters with six meter operation available as a 3/4 wave antenna:

The Stake Stick Vertical

Vertical antennas are great, as are end fed wires. Easy to put up and locate in restricted spaces. But they often put heavy demands on **ground systems and radials**. Here is some of the "bad news" about

grounds, ground enhancements, what does and does not work:

Radial Ground Systems

An entirely new area, Ham antennas for WiFi, extending operation of 802.11b wireless network access to allow high speed digital radio experiments cheaply, even up to and including adding amplifiers which for licensed Ham Radio operators is legal under part 97. When it exceeds part 15 specifications, it should be properly called HSMM or High Speed MultiMedia hamming:

Hinternet Antennas for WiFi and HSMM

For portable operations, many people are attracted to lofting field day type antennas with **gas filled balloons**. Here are details on what it takes to get such a project off the ground:

Balloon Lift With Lighter Than Air Gases

Almost as popular as balloons for portable antenna experiments are **kites**. Here are a few kite related notes:

Kites Suitable for Lifting

The SLOPO is the result of much research with EZnec looking for a reasonable portable antenna that had as little to carry as possible and yet had significant performance. It was hard to beat a simple dipole and come up with a single pole system that had merit, but the SLOPO, the combination of the sloper dipole with a reflecting support pole, making a sort of beam, shows promise:

The SLOPO portable antenna system

The EDZ extended double zepp antenna is a classic. It may have fallen from favor due to its ugly feed point impedance. It is highly reactive at about 120 ohms resistive. But a tuner like a Johnson Matchbox

and parallel line should make short work of that problem, and provide a multiband antenna of merit. It also makes a good gain producing antenna for portable use at the higher bands such as this example at 15 meters:

The 15 Meter Extended Double Zepp

Here are lots of numbers for **wire antennas** on various bands from 160 to 6 meters to help you decide how big a given antenna design will be, what might fit in your *"personal antenna farm"*, and what style of wire antenna might be appropriate for your QTH:

HF Wire Antenna Sizes

General Ham Radio Information

Information on the new **online Ham Radio Licensing** system, how to use it, the new forms and the FCC websites to download the new online forms. This is how all Ham licensing issues will be addressed in the future:

The FCC Universal Licensing System

Camping and Ham Radio, especially QRP or low power, go together well, but here are some things to consider to make life in da tent easier and more fun when the two activities are combined:

Radioactive Camping Tips

Batteries are a great emergency power source, but their ratings and chemistry, especially rechargeable ones, is often *Witchcraft*. Here is some information to help sort them out:

Using Batteries

Jumpstart Powerpacks have appeared that are compact units with a sealed lead acid battery and integral charger. This is a report on one such example and conversion to a highly portable QRP/HT power source:

MightyMite Conversion

If a repeater fails, a Net Control Station will have to establish a **net protocol** on a VHF simplex frequency. Here are some guidelines for both Net Control Stations and Net Participants:

Guidelines for Simplex VHF Net Operations

One of the hardest parts of traffic handling on a net is figuring out how to **count the message**. The message count is an important "checksum" to verify that the entire message was received. Getting it wrong can cause lots of confusion:

Guidelines for Word Counting Message Traffic

This site contains the latest information from the **repeater frequency** coordination database for Hawaii:

Hawaii Repeaters

Salt air and Aluminum or Iron do not mix well. Here is a guide to **goo and gunk** designed to make aluminum connections stay connected and fight the rapid conversion of your new beam to white powder and all steel fasteners to red stains:

"Conductive" Greases

In May 1998 there was a major series of solar events. The result was one of the most intense **HF blackouts and Geomagnetic storms** in a decade. One of the QRP-L list regulars unloaded a massive broadside of information in the form of annotations of solar reports, descriptions, explanations, etc. that constitutes one of the best learning experiences available on solar phenomenon and HF propagation. Suck in your gut, brace yourself and read:

Seven [SOLAR] Days in May

This is an great discussion of working Sporadic E skip on 10 meters from a member of the QRP-L Email list:

Sporadic E on 10 Meters

To help sort out any **RFI** problems you might be having, here is a list of the frequencies of TV channels, both over the air and cable:

Cable, TV, and FM Channel Frequencies

Other Ham Radio Specialty Sites

This is a listing of selected web sites for **Other Hawaii/World/Mainland Ham Clubs of merit**:

Other Ham Club Web Pages

You may find the following Ham radio related web sites useful, they include some very nice **Ham utilities**, some **Callsign lookup** locations, **propagation** related sites:

Other Ham Radio Information Pages

Here are some major **mail order** Ham radio stores, **manufactures** of Ham equipment, and some selected special sites. Most of these we have some personal experience with:

Ham Equipment Suppliers/Manufacturers

These pages are **local Hawaii weather** information pages:

Honolulu NWS and Central Pacific Hurricane Center

University of Hawaii Meteorology

06/02

HFpack. The HF Portable Group. Calling Frequency 18157.5kHz USB or 18158kHz CW

Over 4000 members

What HF frequencies are open now?

- **Hourly Frequency Maps for Band Openings**

  These maps are from actual hourly ionosphere soundings.

- **HFpack Frequency Chart 2004c**

  Use this chart for QSOs with other HF portable operators.

HFpack is an international resource for portable High Frequency communications.

HFpack provides an information exchange about transceivers, antennas, systems, packs, propagation, new developments and techniques in HF portable operation. The HFpack egroup and on-the-air schedules are active with amateur radio around the world.

HFpack membership is free and open to all radio operators.

Join HFpack Free

Read HFpack egroup

News **Features** **On Air Schedule**

HFpack Pedestrian Antenna Shootout Reports

- Main Page
- Site Map
- HFpack Hall of Fame
- Read HFpack E-Group
- On-The-Air Schedule
- Fotos of the Week
- Transceivers
- Antennas
- Packs
- Power Sources
- Bicycle and HPM
- HFpackers Showcase
- Database
- Join HFpack E-Group
- HFpacking Stuff

Our Motto: Don't Just Listen, Call CQ HFpack!

Links

FAQ

Be a part of the exciting HF portable group.
Join HFpack on the air and on the internet.

About HFpack

HFpack is the result of a new type of group developed on-the-air and over the internet. The HFpack e-group is a spam-free moderated discussion group, primarily utilized by amateur radio operators who are interested in HF portable communications and equipment. Subscription and "membership" in the HFpack e-group is free. It is currently active using the Yahoo groups server system, where messages can be read and posted via the web, direct email, or email daily digest modes. Participants in the HFpack e-group observe friendly cooperation and good netiquette. HFpack also maintains HFpack Hall of Fame, the amateur radio Pedestrian Mobile and Human Powered Mobile world record database.

HFpack News

HFpack Frequency Chart 2004c released

Click here for listing

OA9 / KQ6XA Expedition Peru 2004

HF Portable 02 July - 25 August

In addition to the regular daily HFpack schedule at 1630z and 2230z on 18157.5kHz, the expedition will operate on the following schedule:

Time UTC	Meter Band	Frequency kHz	Schedule
00:30z	17m	18157.5 USB or 18158 CW	Daily
01:00z	40m	7087.5 LSB or 7087 CW	Daily
00:45z	15m	21437.5 USB or 21438 CW	As needed
03:50z	30m	10117.5 CW	As needed
xx:xx	17m	18095 CW	QSY Freq for CW

The Cave Expedition Peru 2004 will operate on HF using a 20W transmitter and battery power charged by solar. Radio operators around the world are invited to QSO with OA9/KQ6XA. Please call at the above scheduled times and

Cave Expedition Peru 2004

frequencies on SSB or CW. Listen for response in CW or SSB. QSL via eQSL.			

HFpack.com Features

HFpack Pedestrian Shootout Reports 2001, 2002

HFpack is dedicated to furthering the state of the art for HF Portable. The HFpack Pedestrian Shootouts 2001 and 2002 measured pedestrian antenna systems to a fraction of a decibel, and the reports are presented as a service to radio operators around the world... (more info)

HFprojects! HFpacker Amp, Z-Match Tuner,

Portable Battery Pack - Charger

Virgil K5OOR, an active HFpacker, has developed a new homebrew 35W SSB/CW pedestrian/portable amplifier, Z-Match tuner, and battery pack in small packages. Hundreds of radio operators are now building them together and buying parts together, thereby reducing the cost. A group build is now open. Visit the HFprojects.com website for more information about HF Projects.

HFpack On The Air

HFpack Daily Global Calling:
18157.5 kHz USB or 18158 CW

Day time calling 30 minutes past each hour

HFpack On The Air
Saturday and Sunday Schedule

UTC Time	Freq kHz	Region
16:30z	18157.5 USB or 18158 CW	Global
22:30z	18157.5 USB or 18158 CW	Global
01:45z	5371.5 USB	USA
09:15z	3587.5 LSB	VK/ZL
18:40z	7087.5 LSB	Europe

HFpack Founder
Bonnie KQ6XA

HFpack egroup Moderators
Bonnie KQ6XA
Bob AB7ST
Virgil K5OOR
Ken N0VZ
Budd W3FF

HFpack Steering Committee
Bob AB7ST
Virgil K5OOR
Ken N0VZ
Budd W3FF
Tom G0SBW
Kenji JA1NGA
Sasi 9V1SM
Brad VK2QQ
Bill KR8L
Tom KC5UN
Joe W5SAN
Ken WB6MLC
Mark KI0PF
Sharyl W3VET
Bonnie KQ6XA
Randy K3QO
Bill K6ACJ
John K6ERO
Bill KM4P
Jerry WA2OMU
Jay N1RWY

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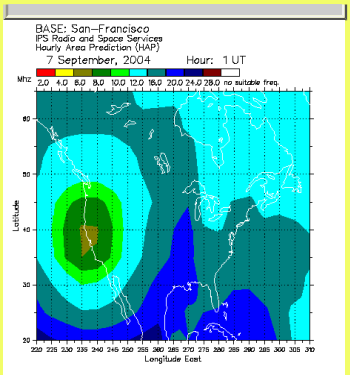

Don't just listen... Call CQ HFpack!

HFpack's special On-The-Air egroup is called HFnow. It is the place to post skeds, operating times and frequencies.

Hourly Propagation Maps

Optimum Frequency Maps Centered on World Cities Based Upon Hourly Ionosphere Soundings

Updated at 40 minutes past the hour; please reload your browser.

North America NVIS - North America Regional

Europe NVIS - Europe Regional

Australia NZ NVIS - Australia NZ Regional

Far East Asia NVIS - Far East Asia Regional

These maps are made from data gathered in real time around the world using HF ionospheric radar systems called ionosondes. Ionosondes bounce HF radio signals off the ionosphere to measure the height of the reflective zones and signal strength vs frequency. The maps are generated using the ionosonde data to make projections based upon a base station working a mobile. To use the maps on this site, locate a map for city near your QTH. Follow the color freqeuncy contours to locate an area you wish to communicate with. For optimum signals, tune your radio to the ham band closest to the MHz of the color range indicated by the map.

Mapping source: Australian Government IPS Radio and Space Services

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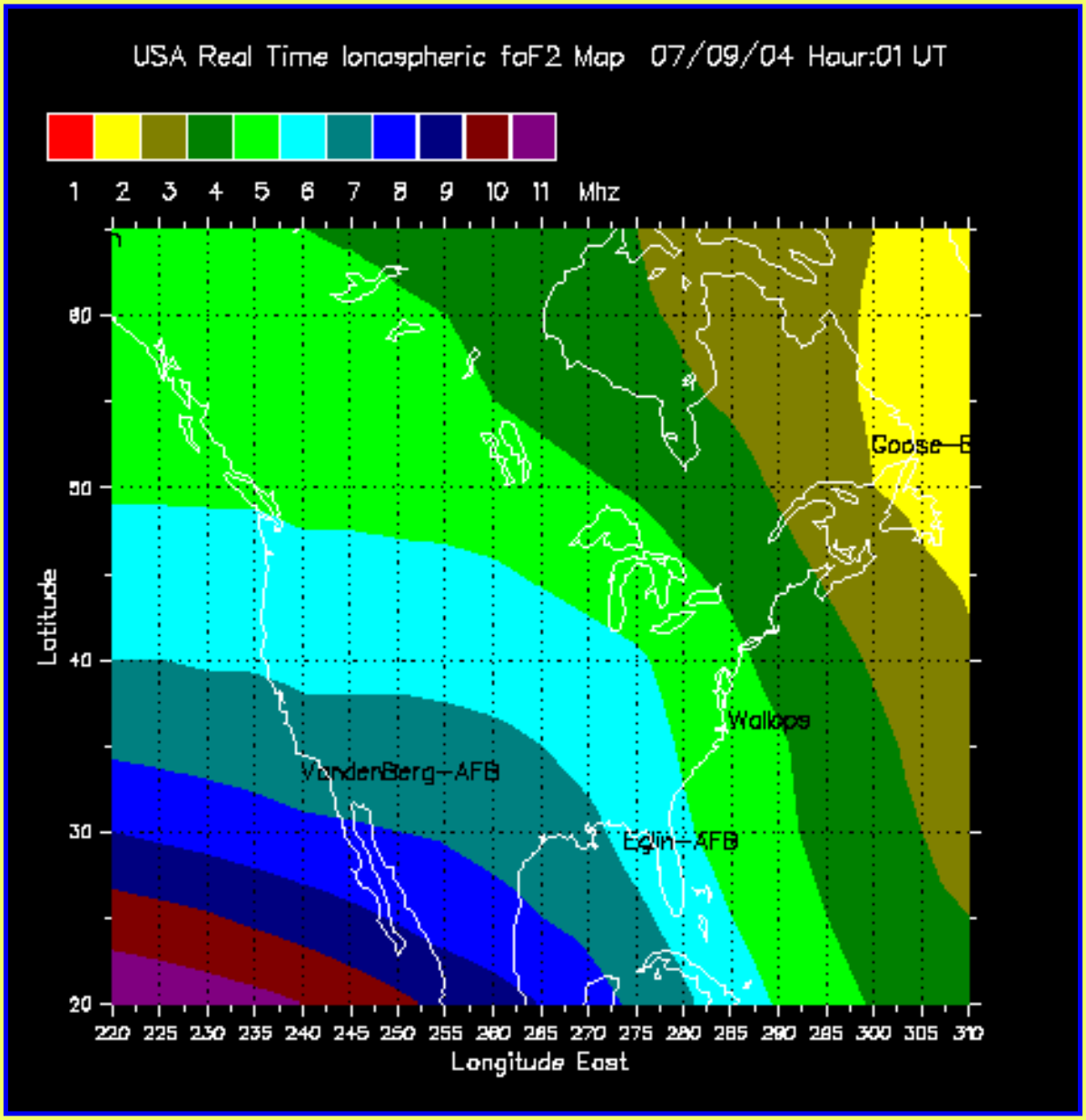
NORTH AMERICA NVIS OPTIMUM FREQUENCIES for communications within 200 miles



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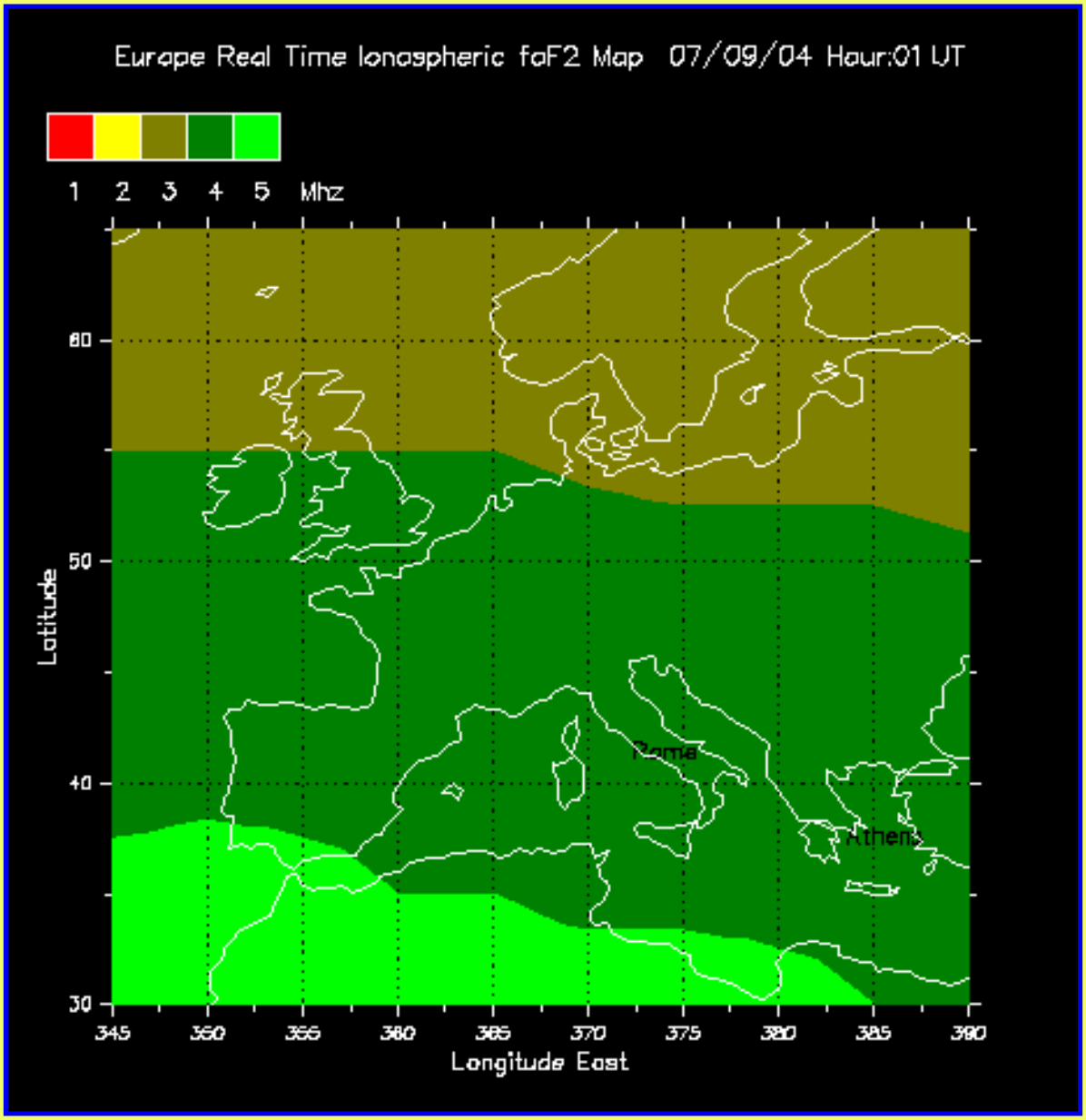
EUROPE NVIS OPTIMUM FREQUENCIES for communications within 300 kilometers



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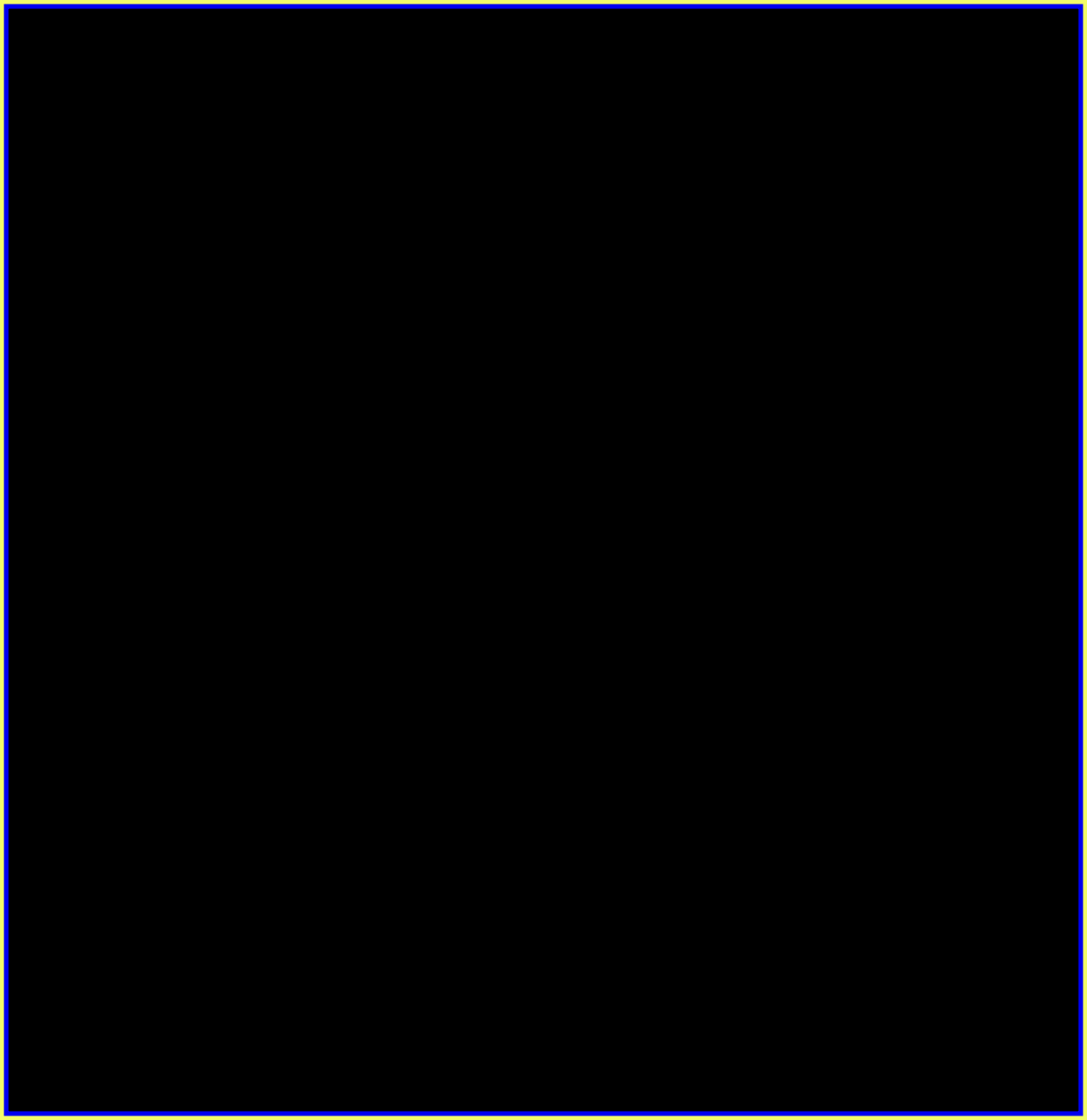
VK/ZL Australia - South Pacific
NVIS OPTIMUM FREQUENCIES
for communications within 300 kilometers



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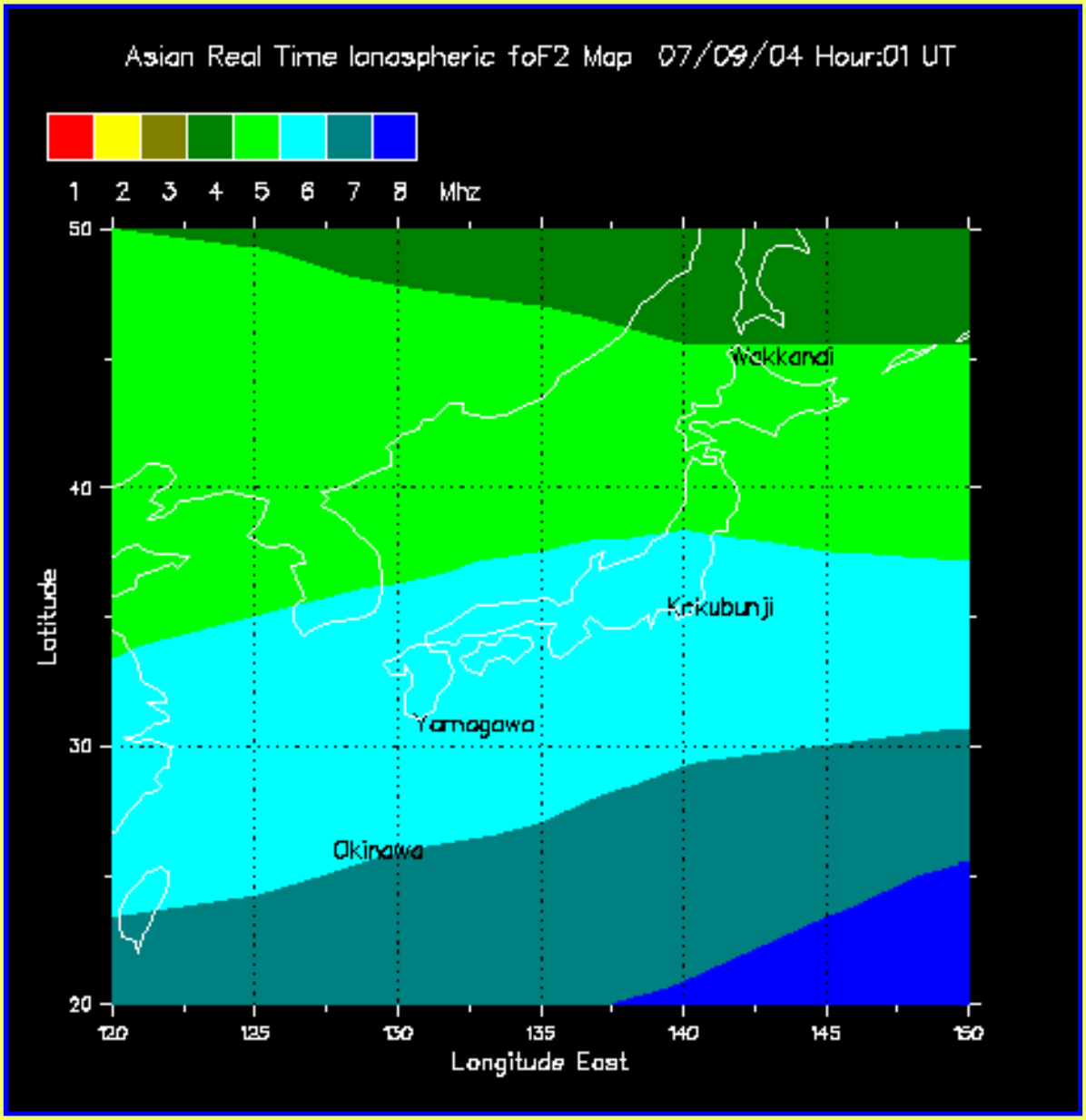
Far East Asia NVIS OPTIMUM FREQUENCIES for communications within 300 kilometers



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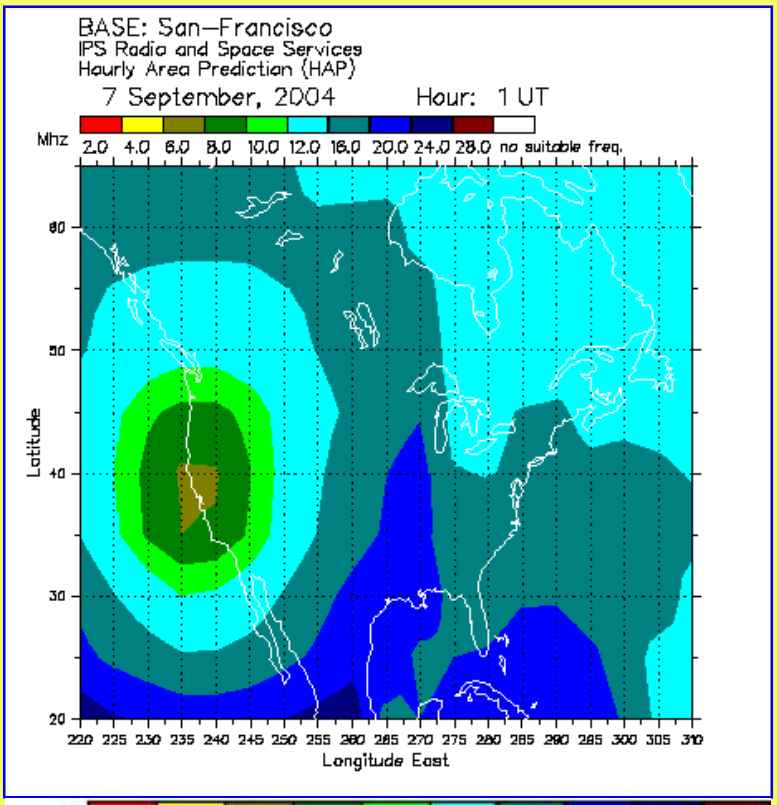
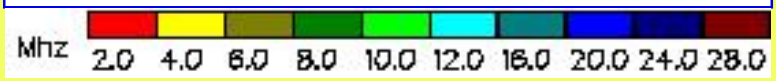
Mapping source: Australian Government IPS Radio and Space Services

Regional Optimum Frequencies

Centered on world cities
click on map to zoom

San Francisco





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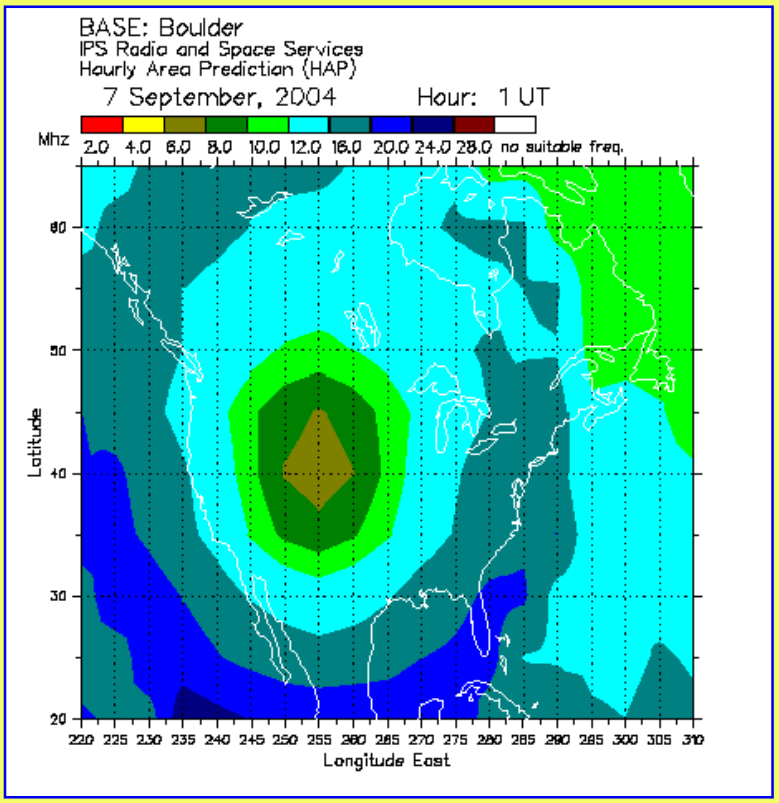

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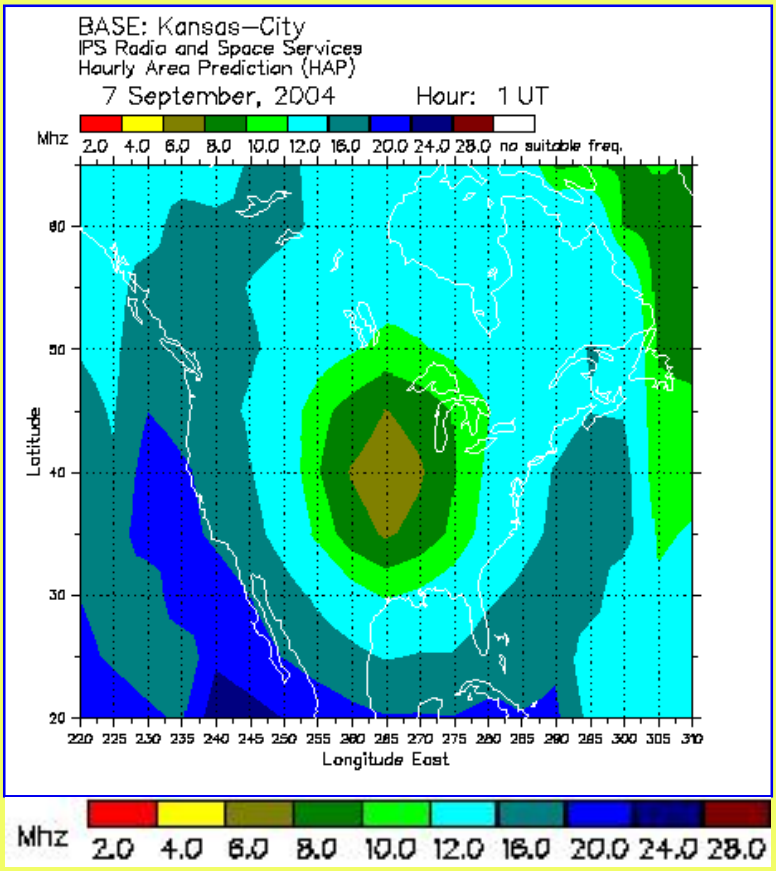

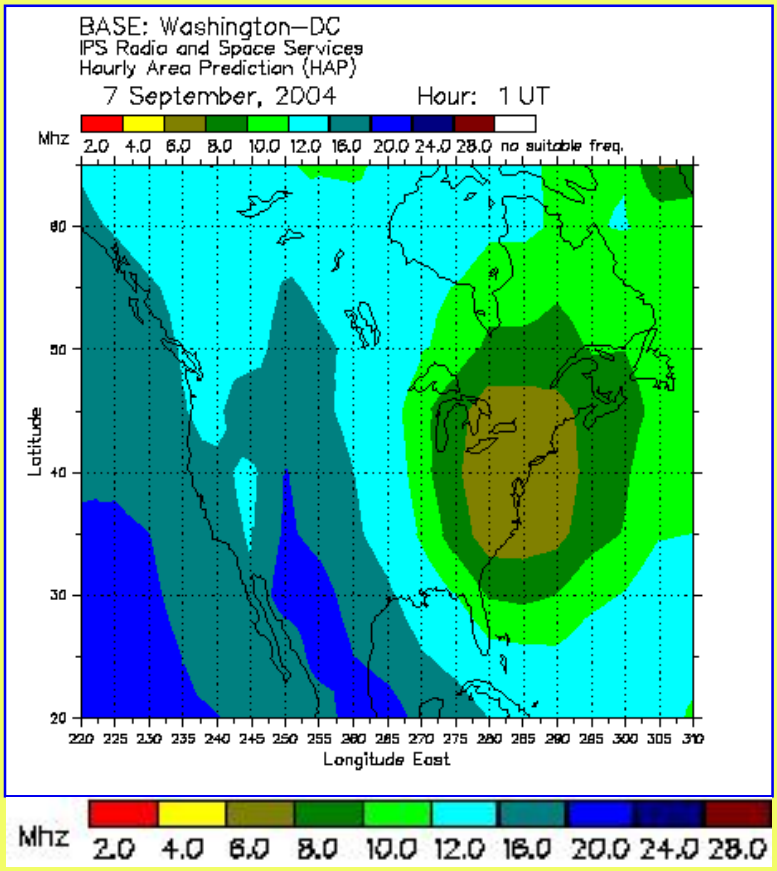

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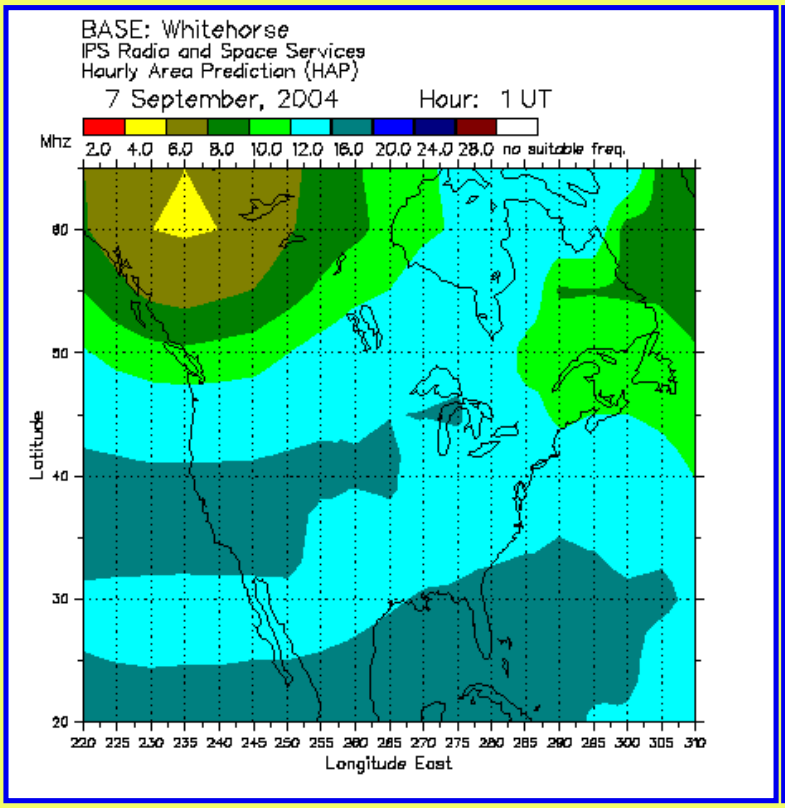

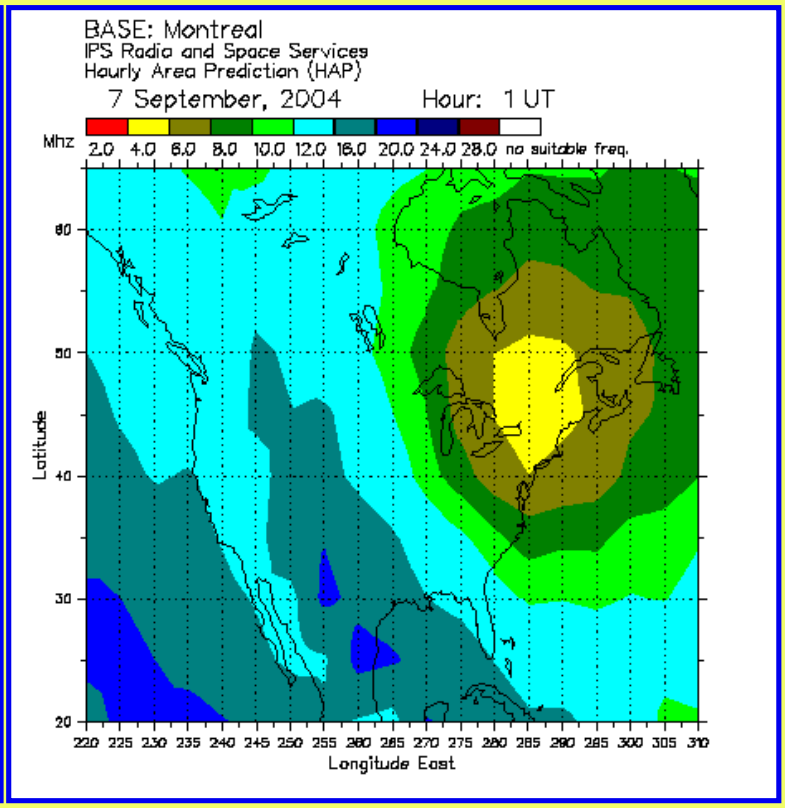

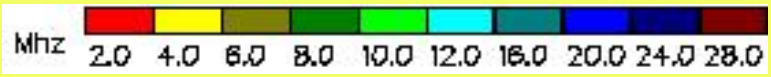

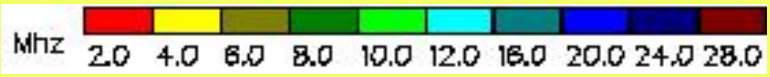

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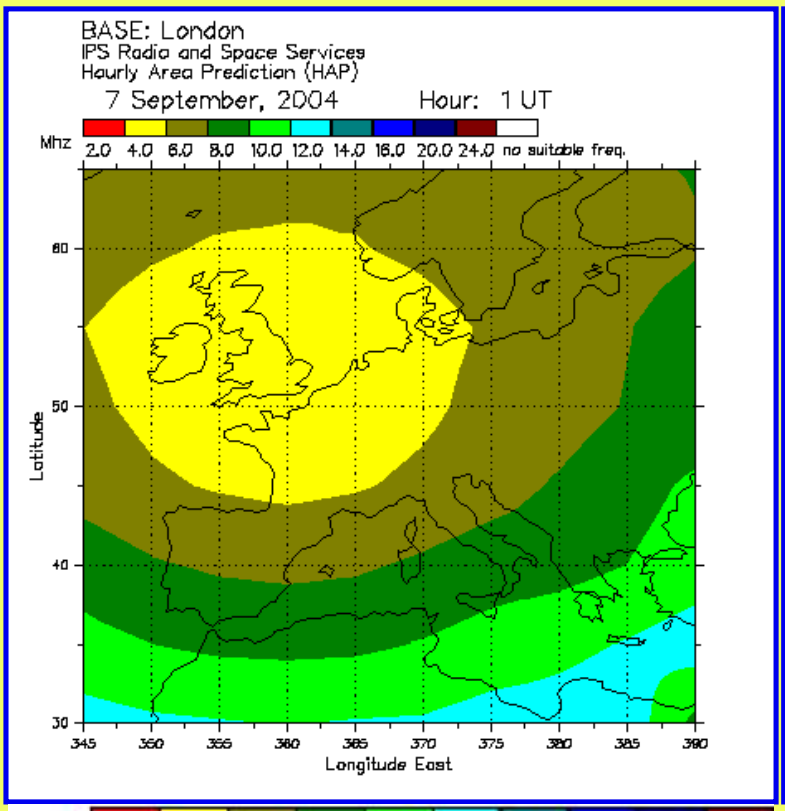

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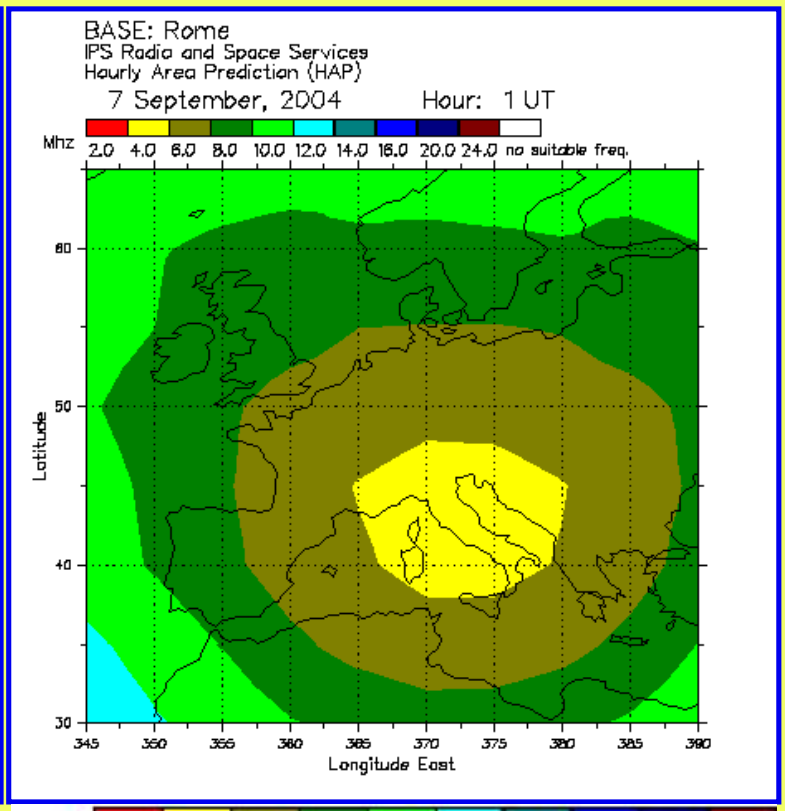

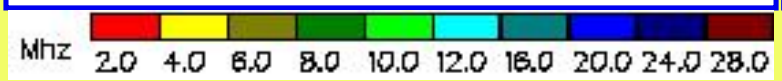

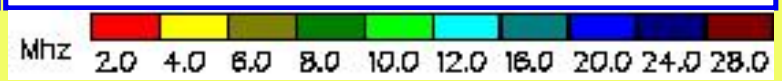

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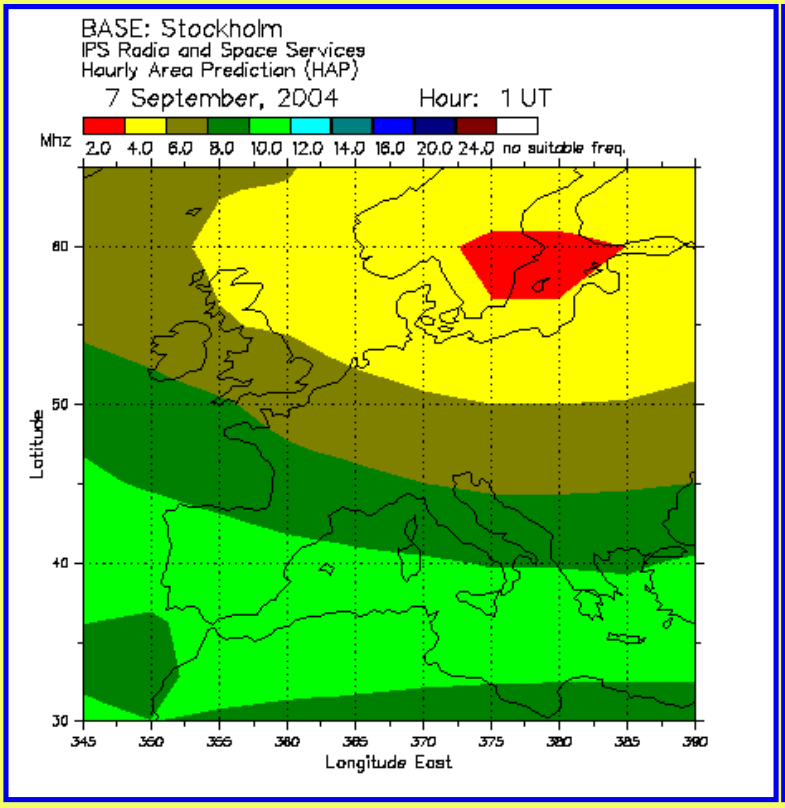

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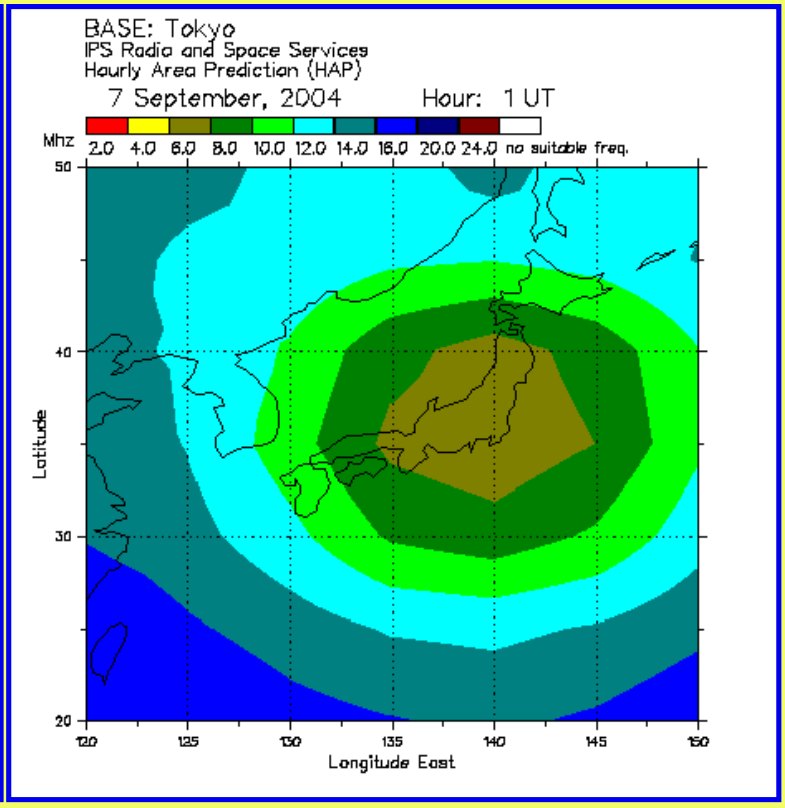

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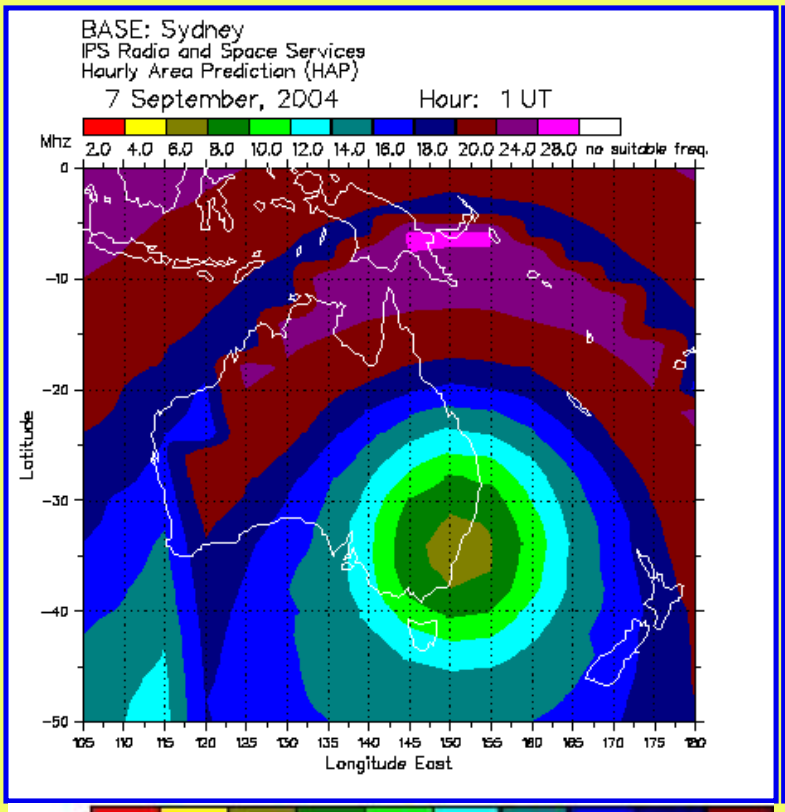

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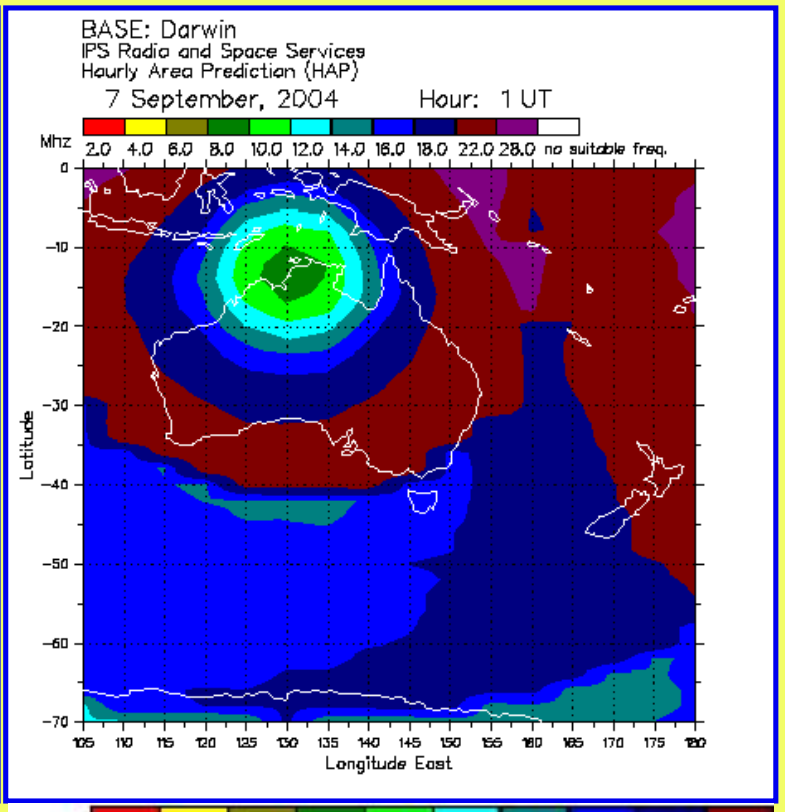

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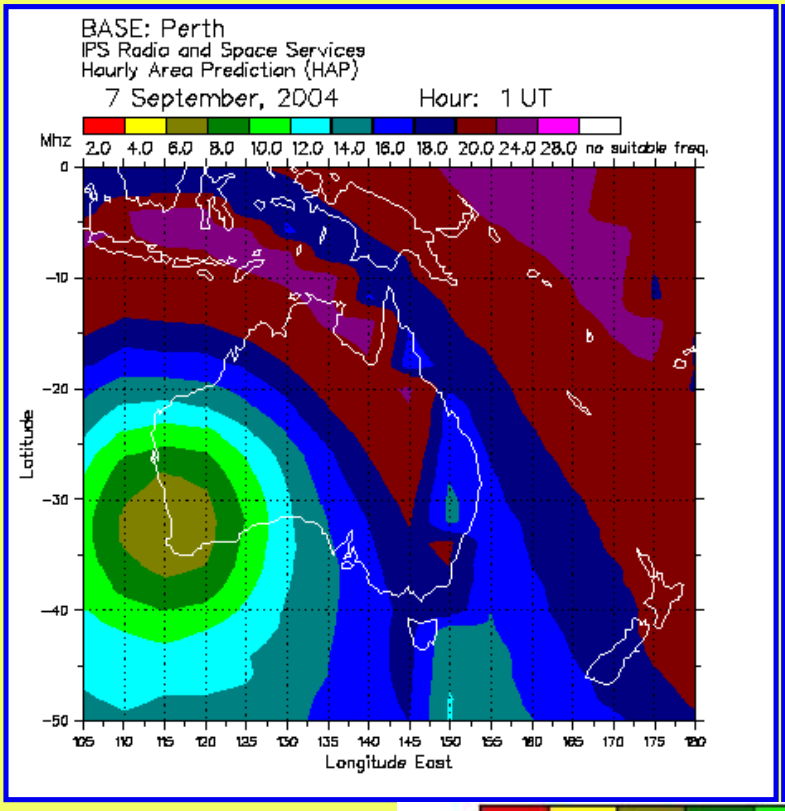

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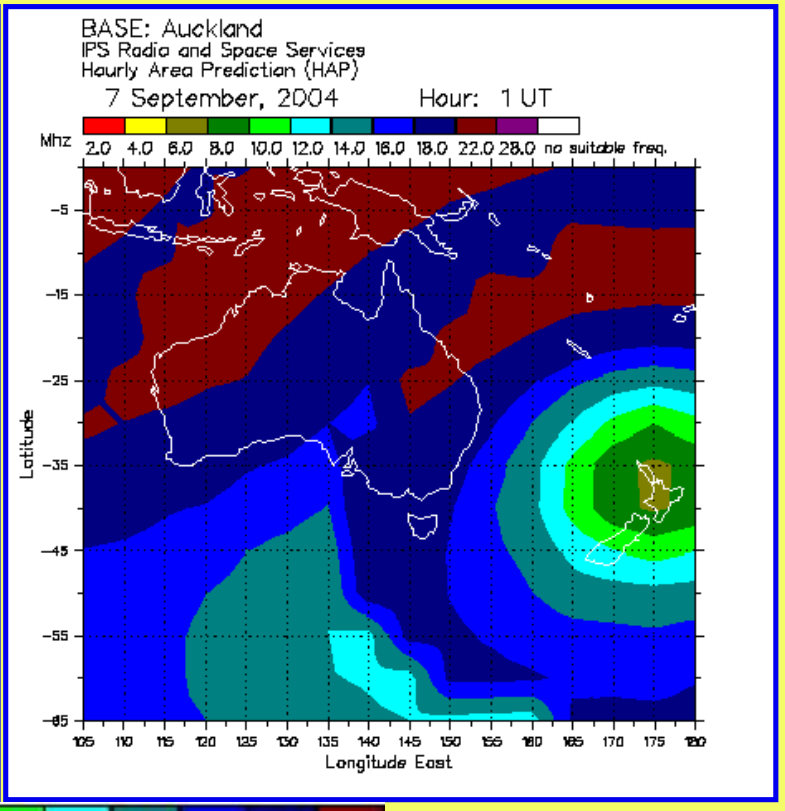

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Rhombic Antenna, 30m-6m

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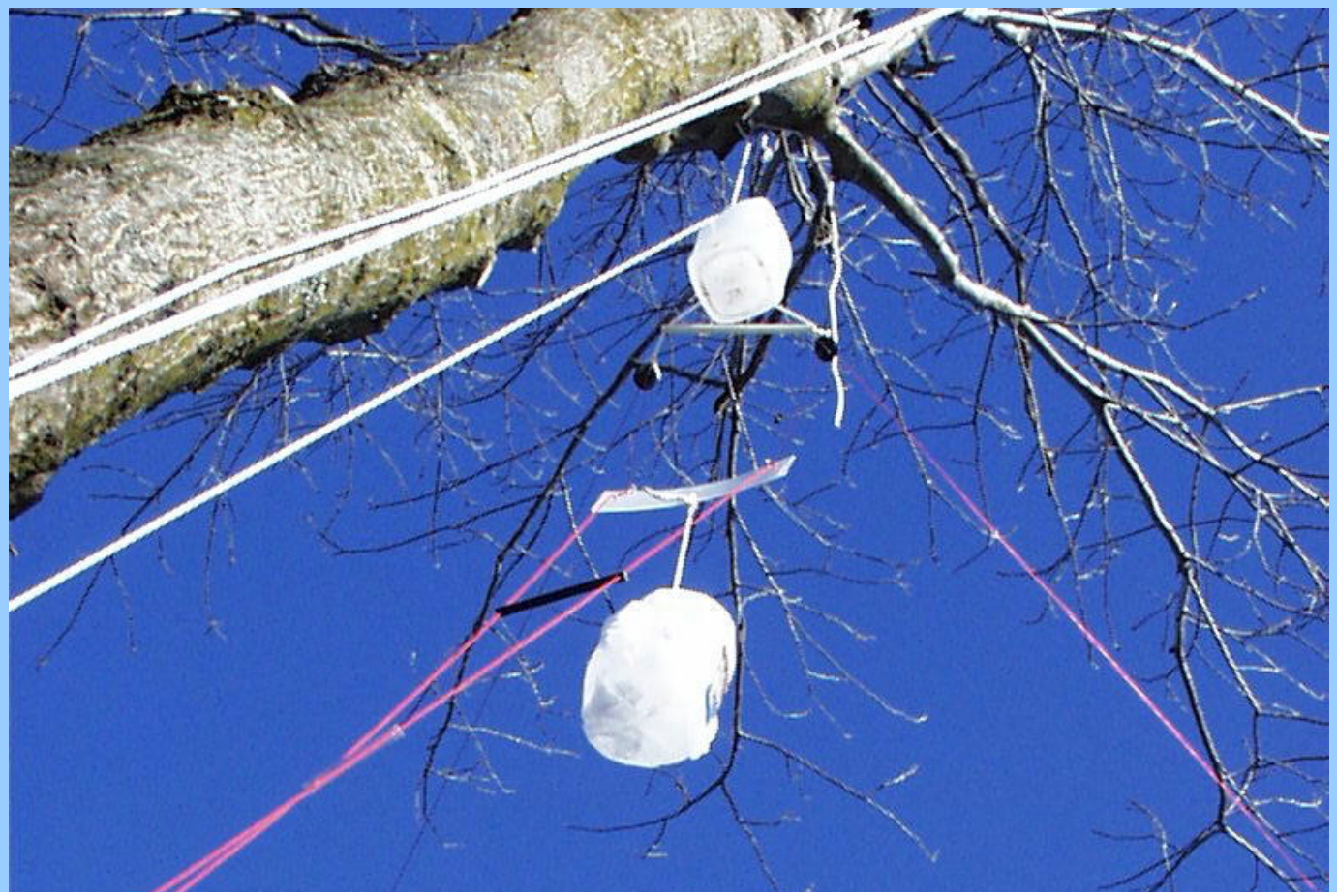
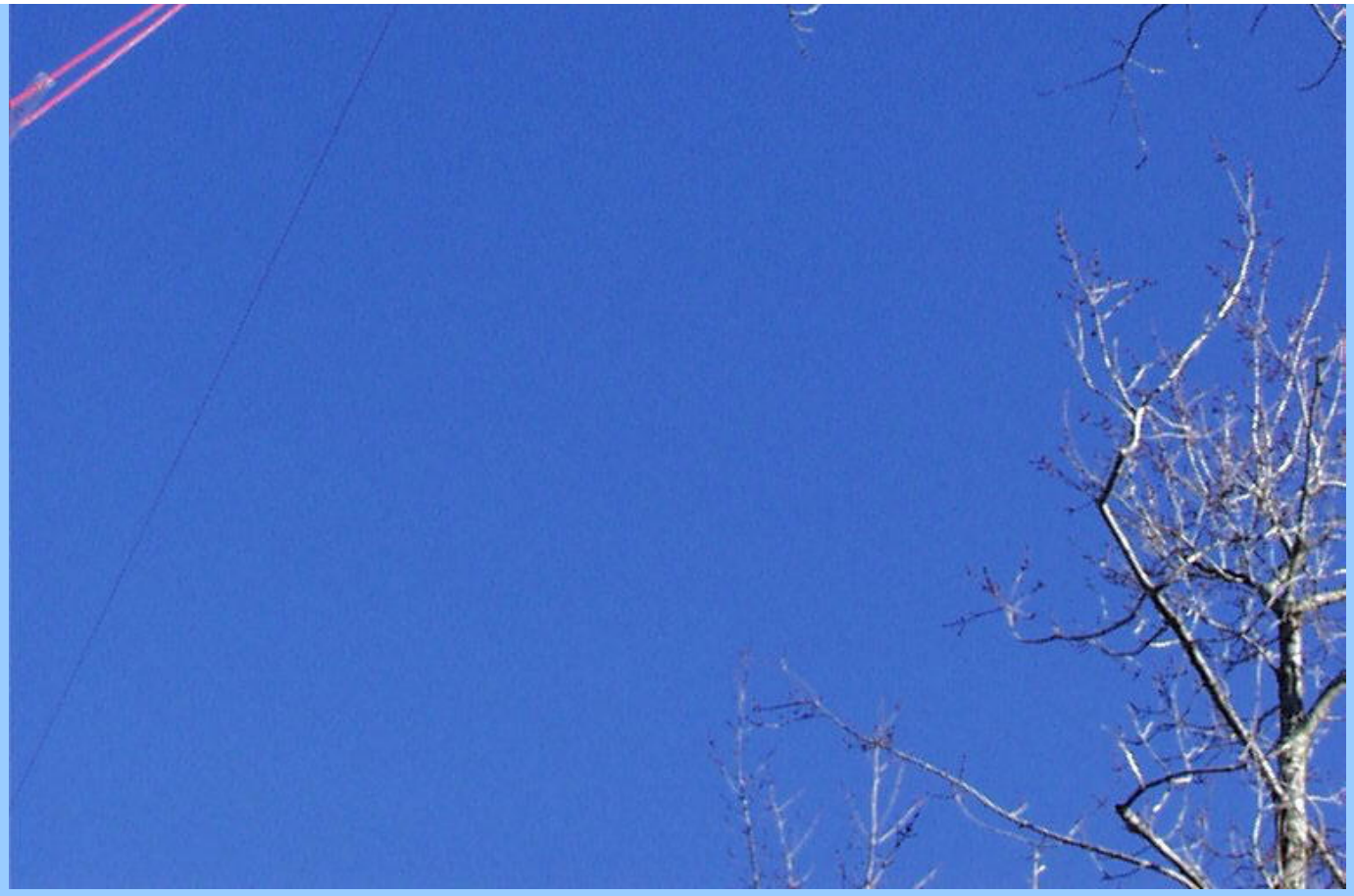
Rhombics are high gain directive wire antennas. A rhombic is shaped like a diamond, if you're looking down on it. I.e., there are 4 'legs', 2 end 'points', and 2 side 'corners'.

This one was designed for 10 meters and 6 meters, but it's proven useful all the way down to 30 meters. Gain figures are:

6m	11db	about equal to a 7el yagi
10m	10db	about equal to a 5el yagi
15m	8db	about equal to a 4el yagi
20m	6db	about equal to a 3el yagi
30m	4db	about equal to a 2el yagi

(gain figures are referenced to a horizontal dipole at the same height)





This particular rhombic is 120' long, 70' on a leg, 25' high, and it's a "terminated" rhombic, which means:

- it is *not* a resonant antenna
- it is *inherently* broadband, 2 octaves or even more
- it does *not* require a tuner...*if* you can match it to the xmitter properly!

Rhombics have a high input impedance of about 600-800 ohms, so it's a non-trivial matter to achieve a match to 50 ohm coax *while still retaining the very broadband nature inherent to the terminated rhombic.*

Matching the Terminated Rhombic:

After much wailing and gnashing of teeth (i.e. experimentation with many different balun transformer designs), I came to the conclusion that the books are right; you just can't get a ferrite balun xformer to work well at both high frequencies *and* high impedance. About 200-300 ohms is the limit for "broadband" work.

This led me to do some research on "tapered lines", the result of which is shown in the pictures here. After several prototypes, I settled on the tapered 2-wire transmission line shown here. I used a modified exponential taper, from 10.00" to .300", over a length of 33', creating a broadband linear transformer from 750 ohms to 200 ohms. The 200 ohm end feeds into a 'normal' 4:1 ferrite balun, which completes the 16:1 ratio required to match to 50 ohm coax.

Results with Tapered Line:

With this setup, I've finally achieved a good SWR over the entire bandwidth of the antenna; less than 1.3:1 from 10-60 Mc. At the important (to me) 15, 10, and 6m bands, the SWR is under 1.2:1. I was never able to get less than 2.5:1 when using two 4:1 baluns in series, so to me this represents a big success.

Power Handling:

Tapered line also has the interesting property, unlike ferrite xformers, of being able to handle lots and lots of power....<g> Now I need to work on a higher power terminator; and then I'll be able to run an amp on the rhombic if I choose. The current terminator is rated 50w continuous; enough for a 100w transceiver, but not for a linear.

Mechanical Design:

The picture above is a shot looking up the feed-end tree; showing the tapered line and the wire-supports. The entire antenna "floats" in its supports. I.e., it is only firmly attached down at the terminator end. At every other corner, the wires slide freely through their "pulleys", which are poly egg-insulators. This design automatically equalizes tension on all legs, and has proven to work very well. The antenna is always flat, and it handles high winds and tree-branch hits with ease.

The rhombic itself is made from #18 enameled copper wire, and the tapered-line is made from #18 stranded tinned-copper PVC insulated wire.

The 1-gallon jugs are the counterweights used to tension the antenna. The amount of water was adjusted to produce the correct tension in the wire.

It came as an unpleasant surprise to me that it's very difficult to take good pictures of thin wires against the sky and trees! <grin> In any case, below is another shot of the tapered line (red vertical wires). You can see some of the "spreaders", which were cut from IC-tubes.

The extra ropes in the picture are the halyards for the ends of the 80m and 40m dipoles, which also use this tree for support.

I'm going to try re-shooting these pictures when the sky is a 'white' overcast, in hopes that the wires will be more visible. I also hope to get a better focus. Usually this Oly D500-L shoots outstandingly sharp pics. Maybe I was vibrating...<g> Stay tuned...

It took a *lot* of clearing to put up this antenna! Three hundred and fifty feet of this....but hey, we needed the firewood anyway <g>.

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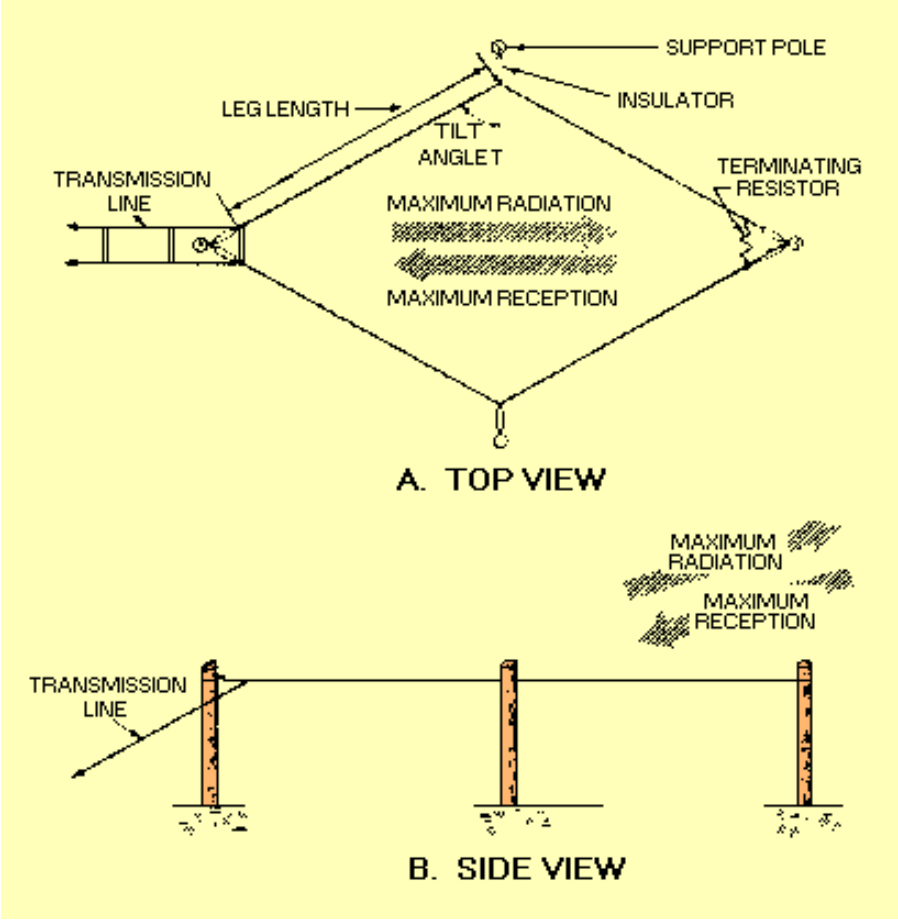
RHOMBIC ANTENNA

The highest development of the long-wire antenna is the RHOMBIC ANTENNA (see figure 4-37). It consists of four conductors joined to form a rhombus, or diamond shape. The antenna is placed end to end and terminated by a noninductive resistor to produce a uni-directional pattern. A rhombic antenna can be made of two obtuse-angle V antennas that are placed side by side, erected in a horizontal plane, and terminated so the antenna is nonresonant and unidirectional.

Figure 4-37. - Basic rhombic antenna.



The rhombic antenna is WIDELY used for long-distance, high-frequency transmission and reception. It is one of the most popular fixed-station antennas because it is very useful in point-to-point communications.

Advantages

The rhombic antenna is useful over a wide frequency range. Although some changes in gain, directivity, and characteristic impedance do occur with a change in operating frequency, these changes are small enough to be neglected.

The rhombic antenna is much easier to construct and maintain than other antennas of comparable gain and directivity. Only four supporting poles of common heights from 15 to 20 meters are needed for the antenna.

The rhombic antenna also has the advantage of being noncritical as far as operation and adjustment are concerned. This is because of the broad frequency characteristics of the antenna.

Still another advantage is that the voltages present on the antenna are much lower than those produced by the same input power on a resonant antenna. This is particularly important when high transmitter powers are used or when high-altitude operation is required.

Disadvantages

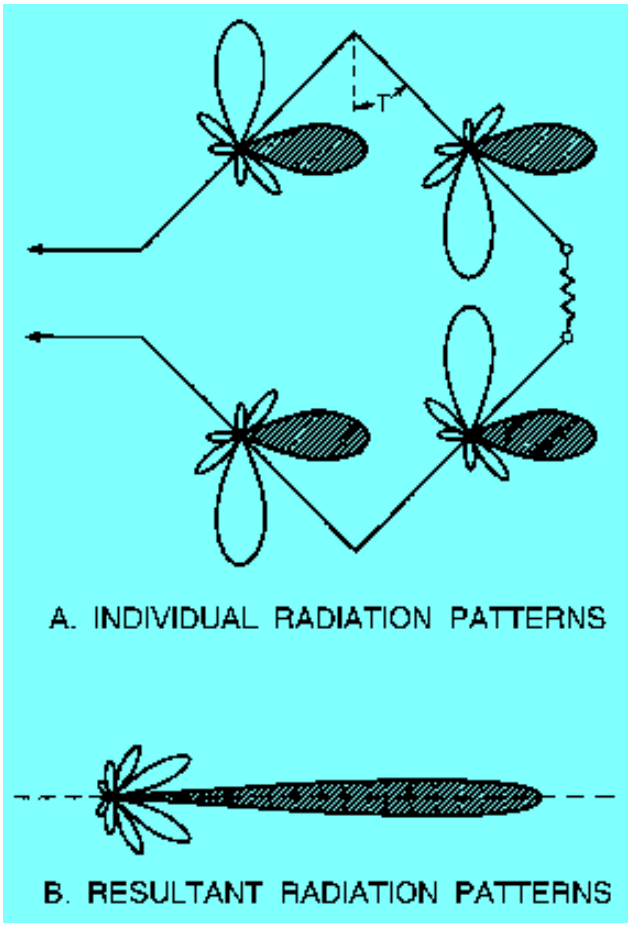
The rhombic antenna is not without its disadvantages. The principal one is that a fairly large antenna site is required for its erection. Each leg is made at least 1 or 2 wavelengths long at the lowest operating frequency. When increased gain and directivity are required, legs of from 8 to 12 wavelengths are used. These requirements mean that high-frequency rhombic antennas have wires of several hundred feet in length. Therefore, they are used only when a large plot of land is available.

Another disadvantage is that the horizontal and vertical patterns depend on each other. If a rhombic antenna is made to have a narrow horizontal beam, the beam is also lower in the vertical direction. Therefore, obtaining high vertical-angle radiation is impossible except with a very broad horizontal pattern and low gain. Rhombic antennas are used, however, for long-distance skywave coverage at the high frequencies. Under these conditions low vertical angles of radiation (less than 20 degrees) are desirable. With the rhombic antenna, a considerable amount of the input power is dissipated uselessly in the terminating resistor. However, this resistor is necessary to make the antenna unidirectional. The great gain of the antenna more than makes up for this loss.

Radiation Patterns

Figure 4-38 shows the individual radiation patterns produced by the four legs of the rhombic antenna and the resultant radiation pattern. The principle of operation is the same as for the V and the half-rhombic antennas.

Figure 4-38. - Formation of a rhombic antenna beam.



Terminating Resistor

The terminating resistor plays an important part in the operation of the rhombic antenna. Upon it depend the unidirectivity of the antenna and the lack of resonance effects. An antenna should be properly terminated so it will have a constant impedance at its input. Terminating the antenna properly will also allow it to be operated over a wide frequency range without the necessity for changing the coupling adjustments at the transmitter. Discrimination against signals coming from the rear is of great importance for reception. The reduction of back radiation is perhaps of lesser importance for transmission. When an antenna is terminated with resistance, the energy that would be radiated backward is absorbed in the resistor.

Q.47 What is the main disadvantage of the rhombic antenna? **Answer**

TURNSTILE ANTENNA

The TURNSTILE ANTENNA is one of the many types that has been developed primarily for omnidirectional vhf communications. The basic turnstile consists of two horizontal half-wave antennas mounted at right angles to each other in the same horizontal plane. When these two antennas are excited with equal currents 90 degrees out of phase, the typical figure-eight patterns of the two antennas merge to produce the nearly circular pattern shown in figure 4-39, view A. Pairs of such antennas are frequently stacked, as shown in figure 4-40. Each pair is called a BAY. In figure 4-40 two bays are used and are spaced 1/2 wavelength apart, and the corresponding elements are excited in phase. These conditions cause a part of the vertical radiation from each bay to cancel that of the other bay. This results in a decrease in energy radiated at high vertical angles and increases the energy radiated in the horizontal plane. Stacking a number of bays can alter the vertical radiation pattern,

causing a substantial gain in a horizontal direction without altering the overall horizontal directivity pattern. Figure 4-39, view B, compares the circular vertical radiation pattern of a single-bay turnstile with the sharp pattern of a four-bay turnstile array. A three-dimensional radiation pattern of a four-bay turnstile antenna is shown in figure 4-39, view C.

Figure 4-39. - Turnstile antenna radiation pattern.

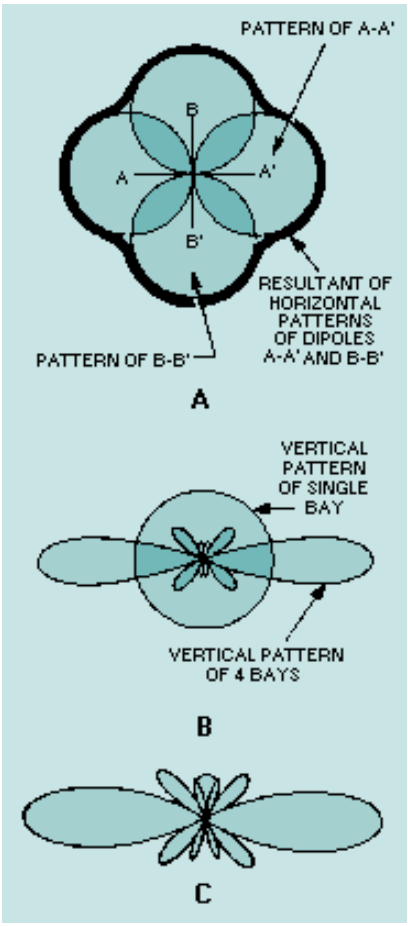
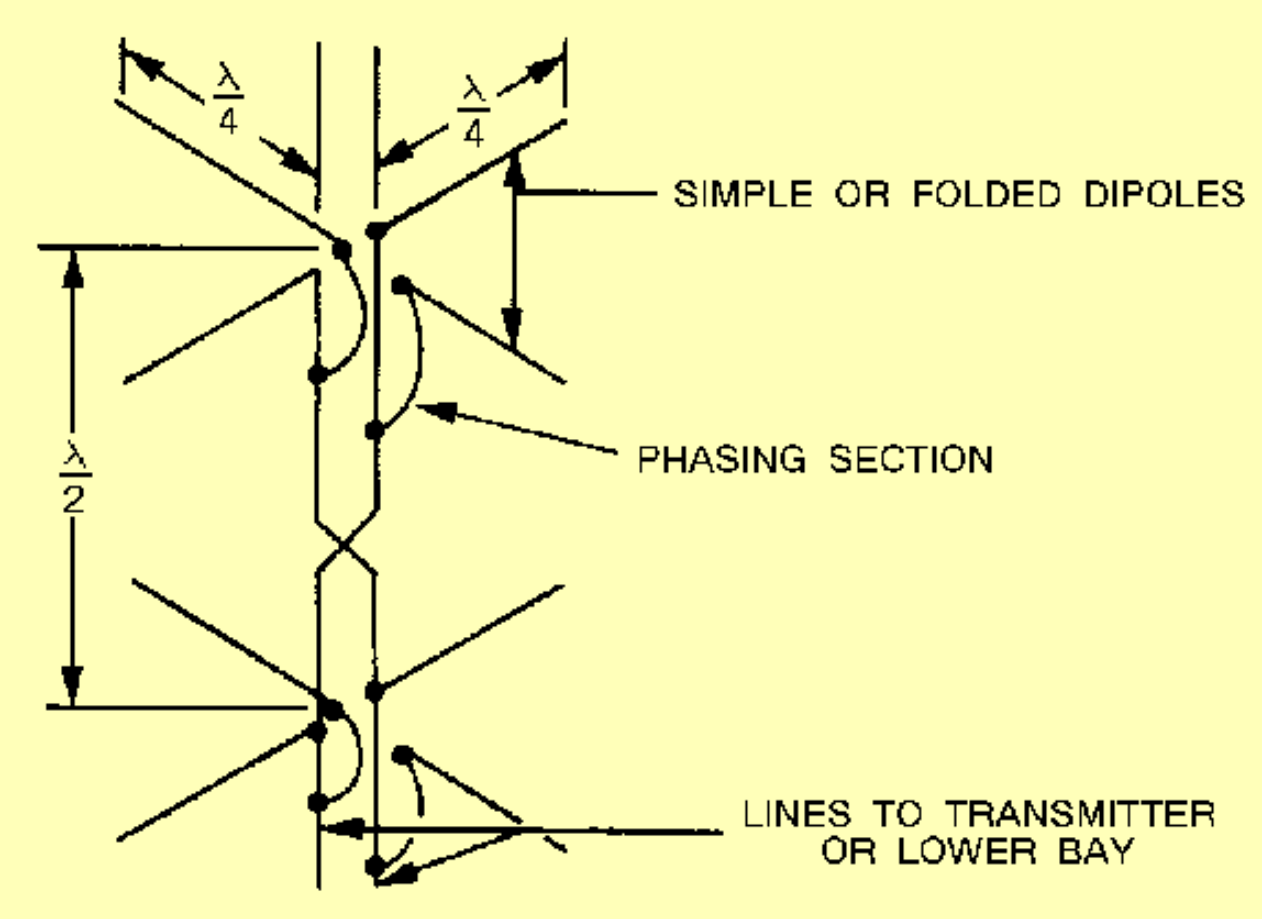


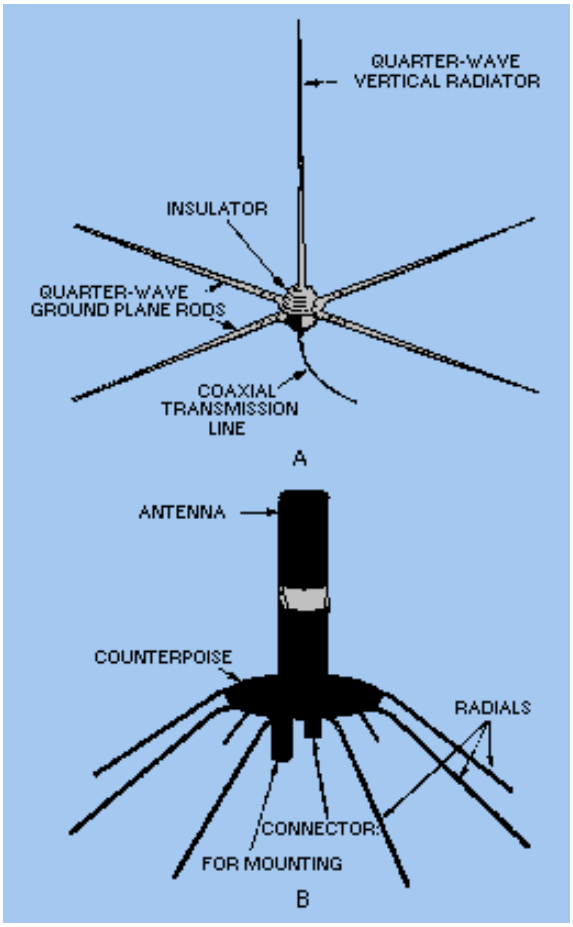
Figure 4-40. - Stacked turnstile antennas.



GROUND-PLANE ANTENNA

A vertical quarter-wave antenna several wavelengths above ground produces a high angle of radiation that is very undesirable at vhf and uhf frequencies. The most common means of producing a low angle of radiation from such an antenna is to work the radiator against a simulated ground called a GROUND PLANE. A simulated ground may be made from a large metal sheet or several wires or rods radiating from the base of the radiator. An antenna so constructed is known as a GROUND-PLANE ANTENNA. Two ground-plane antennas are shown in figure 4-41, views A and B.

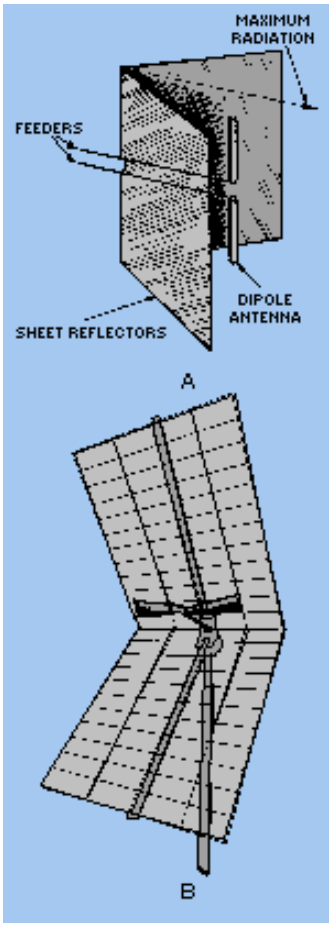
Figure 4-41. - Ground-plane antennas.



CORNER REFLECTOR

When a unidirectional radiation pattern is desired, it can be obtained by the use of a corner reflector with a half-wave dipole. A CORNER-REFLECTOR ANTENNA is a half-wave radiator with a reflector. The reflector consists of two flat metal surfaces meeting at an angle immediately behind the radiator. In other words, the radiator is set in the plane of a line bisecting the corner angle formed by the reflector sheets. The construction of a corner reflector is shown in figure 4-42. Corner-reflector antennas are mounted with the radiator and the reflector in the horizontal position when horizontal polarization is desired. In such cases the radiation pattern is very narrow in the vertical plane, with maximum signal being radiated in line with the bisector of the corner angle. The directivity in the horizontal plane is approximately the same as for any half-wave radiator having a single-rod type reflector behind it. If the antenna is mounted with the radiator and the corner reflector in the vertical position, as shown in view A, maximum radiation is produced in a very narrow horizontal beam. Radiation in a vertical plane will be the same as for a similar radiator with a single-rod type reflector behind it.

Figure 4-42. - Corner-reflector antennas.



Q48. What is the primary reason for the development of the turnstile antenna? **Answer**

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Rhombic Antenna Home Page

While amateur radio operators today largely employ rotating array antennas such as the Yagi dipole array and the cubical quad antenna, some still use wire antennas, some of which can afford higher gain and lower receive noise characteristics. In fact, the rhombic antenna is usually described as the "King of antennas" because of it's very desireable characteristics. The rhombic antenna is basically a diamond-shaped wire curtain that is made of 4 wires, each several wavelengths long connected to form a "diamond" or rhombus shape . The diamond is constructed with the narrowest ends left open for the feed point on one end and a non-inductive resistive termination on the other. This creates the terminated (also called non-resonant) rhombic which is a unidirectional antenna with broad bandwidth. Non-terminated (also called resonant) rhombics do work but have narrower useful bandwidth and are bidirectional (non-resonant rhombics are fed on one end and the opposite end is left open. In my experience both non-resonant and resonant rhombics work quite well. All wire antennas tend to be less "noisy" on receive and longer antennas like the rhombic are very low noise antennas with the solitary exception of precipitation static (snow and rain add small charges to the curtain and you get very high noise under such conditions). Precipitation static can be minimized by using termination resistors with a grounded center tap (leaks static to ground continuously).

The sides of the rhombic (with the broadest included angle) are constructed of differing dimensions to determine what radiation angle will be obtained. Half of the angle included by the rhombic side is called the "tilt angle" or theta ø and as the tilt angle is varied, the radiation angle will also vary. There exist various ways to calculate the ideal tilt angle to choose relative to the radiation angle and gain desired for a given rhombic application. Generally the tilt angle is between 65 and 75 degrees. The apex angle is the angle included by the feed point end of the rhombic and is between 30 and 50 degrees and varies with the tilt angle (the sum of half the apex angle and the tilt angle will be 90).

Height is very important for rhombics, as it is for all antennas. Basically, the radiation angle of the main beam of any antenna is related to the height above the ground and the ground characteristics. Generally, for any antenna the following height (in wavelengths)/wave angle (degrees) relationships hold: 0.5/30; 1.0/15; 1.5/8; 2.0/6; 3.0/5. As you can see, to take advantage of the rhombic's extraordinary gain for distant (low angle) communications, you will need to have the antenna at least 1.5 wavelengths high: on 14.0 Mhz, this is around 90 feet, the "usual" height used for rhombic antennas (most rhombics used are in amateur radio usually optimized for 14 Mhz or 20 meter band).

Rhombic antennas are generally fed with open feedlines of around 400 to 600 ohm impedance. This is not critical and any issues related to feeder loss and/or standing wave ratios pales in significance in relationship to this antenna's gain. In the real world things work fine without worry related to feedlines so long as you use a good antenna tuner. Use of baluns is probably unwise as this is likely to limit the useful frequency range of this intrinsically broadbanded antenna. Many commercial applications of the rhombic use what are called "exponential feeders" with the nominal impedance being 300 ohms at the feeder end and with the paired feedlines gradually spreading to a spacing of around 12 inches at the rhombic feedpoint (a distance calculated to provide the rhombic feedpoint impedance of 600-800 ohms). This exponential feeder provides a smooth transition of impedance and acts as a sort of impedance transformer.

In current antenna publications it seems very hard to find anything other than a very abbreviated discussion of rhombics and perhaps some basic design charts and dimensions for a simple rhombic. Little discussion is generally available of the more complex details important to anyone seriously interested in constructing a rhombic antenna. With this in mind, this page has some images of real-world installations and discussion of most of the important practical details you would need to be aware of to construct and maintain a proper rhombic. I am always interested in hearing from anyone with experience with building, maintaining or using rhombics. Images, plan sets and even parts from dismantled rhombics are of interest to me. I will add anything to this page that fits, and then some.

Click here for an abbreviated discussion of practical rhombic antenna design

Click here for a detailed text outlining apects of practical rhombic antenna design (under construction)

Click here to view construction images for the North American Center for Emergency Communications (NACEC) rhombic station in Minnesota

Click here to read about the World War II rhombic station in Scituate Rhode Island (Chopmist Hill)

The three images included here are from the WJCR short wave station in Millersville Kentucky and the images were kindly provided courtesy of WJCR. Note the feeders are attached directly to the antenna with little concern for matching. This is not unreasonable as the loss involved is trivial compared to the gain of the antenna (I feed mine directly as well). Also note the use of three wire curtains (slight increase in gain and smoother impedance characteristics across the broad frequency coverage of the rhombic.

The issue of frequency coverage itself is also worth further discussion. While the rhombic is very broad banded, especially the non-resonant terminated antenna, in my hands when optimized for 14Mhz, the antenna is still useful from 10 through 18 Mhz. On the higher bands (18-30 Mhz), the main beam becomes unaccepably "split" with less effective gain in the resulting lobes in the azimuth plane. The antenna is still useful but limited. Perhaps the answer is to build the larger rhombics with smaller antennas that are optimized for the higher frequencies supported at a lower height from the 4 support poles. The rhombics work well when "nested" in this manner and you can conserve space in this way. Additionally, if you get really carried away, you can use the side poles as ends for other rhombics and vica versa. You can even "stack" rhombics running in different directions with clearances of anything greater than 5 feet on HF. Yes, there is a slight degradation but big rhombics are such good antennas that in the real world you shouldn't care. Designs have been built where rhombics are phased together so that in a given direction, one or more rhombics are fed in phase.

Monsterous rhombic arrays have been built in this manner and one can only guess what kind of signals resulted (it's actually rather frightening)!

Harper, of Bell Laboratories, did much of the pioneering work on the rhombic antenna and determined a famous curve that is largely used to design rhombics. Usually some optimization is needed to make sure that the main lobe and the radiation angle coincide. This usually results in a slight loss of gain because of the slightly shorter legs used for a given angle on the sides. Some additional advantage is obtained by having several wires per leg (several curtains), with up to 2 db more gain resulting.

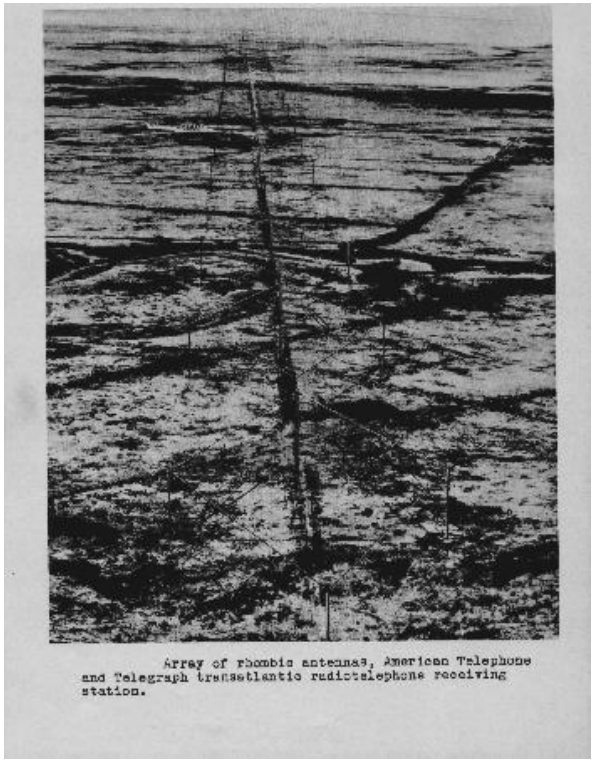
The ability of the rhombic to provide high gain (typically over 9 db for most antennas made and often up into the 20 to 30 db area in arrays of rhombics) , narrow predictable patterns in the azimuth plane, optimized patterns for a given wave angle and physically sturdy construction make the rhombic still the choice antenna for point to point communications operations. Many large rhombic stations existed until recent times and several still are in operation.

Some famous stations included WCC in Chatham Massachusetts, the Rocky Point RCA station and the Palos Verdes Press Wireless Station of the late Don Wallace W6AM's (below). These stations began as commercial stations and had many rhombics, usually with each rhombic covering a given city (again, because you can optimize the wave angle and the azimuth direction is known, you would choose a rhombic with a higher wave angle for cities that are closer and a lower angle for farther destinations). These antennas almost guaranteed an open circuit to the desired locations despite the conditions. Such arrays would usually be fed via open feeders (oftern 4 wire feed lines to limit loss over the thousands of feet of feeders) and if both directions were desired a switching array would be used to select the direction. Stations would usually have a matrix of open frame RF relays in the shack or in a switch house with a DC switch at the operator's station to select the direction desired. The feeders would arrive in the switch area in layers.

The military has used rhombic antennas for years. In fact, in World War II, the War Department had a rhombic antenna kit (TM11-2611) . This manual was designed to allow the construction of rhombics in the field with indiginous timber for supports and using tables to determine the optimum leg length and tilt angle for communications back to the United States. This design is near optimum for the largest rhombic feasible for practical HF communications.

During World War II not far from my station's location was a classified HF listening station where it is said that rhombics there were used to eavesdrop upon German submarines during the war. This station was said to be involved in the breaking of the German Enigma codes during World War II. This station was located on the top of Chopmist Hill, not far from the highest spot in Rhode Island. This location in the town of North Scituate was designed by Thomas Cave of the Federal Communication Commission's Radio Intelligence Division (RID). The RID had leased the 183 acre Stoddard Farm, chosen for it's location and open acreage that was amenable to a large wire array receiving station far from industrial and power line noise sources. Nationwide there were 150 such RID stations used to intercept wartime radio traffic. The Chopmist

Hill station had 85,000 feet of wire antenna curtains aloft. The poles were typically 80 feet high and were still aloft in 1968. Not being rotary antennas caused some problems with the strategic listening needs. During the critical Second World War battle of El Alamein in 1942, Cave was calling the Narragansett Electric Company several times daily to reposition antenna poles to optimize reception (truely the "armstrong rotor" method)! On another occasion the rhombics on Chopmist saved the Queen Mary carrying 10,000 American troops overseas by interceptin orders to sink her sent to the U-boat fleet in the Atlantic. The station was also credited to accurately pinpointing the location of every Axis spy within the US by triangulation with other RID stations. Curiously, Frank W3LPL (who had seen the site in the 60's) has told me that the site had a Swastika-shaped fireplace, I suspect as a reminder of their mission, although I am not certain of this. This site was later proposed as the potential site for the United Nations but was later passed up for New York.

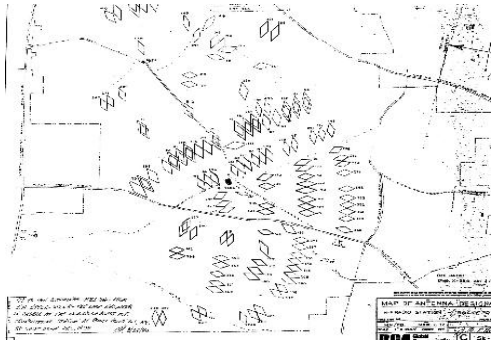


Array of rhombic antennas, American Telephone and Telegraph transatlantic radiotelephone receiving station.

Various commercial rhombic stations exist even today. One is on Cape Cod (Chatham) and is the receiving end of an HF relay link. One of the largest rhombic stations ever was on Long Island at Rocky Point and was an RCA station.

This station was off Route 25A. This station was operational until 1978 and at it's peak had 137 rhombic antennas, many phased together. I have many insulators from this station, including some very distinctive custom-built 4-wire feedline insulators with brass/ceramic turning and termination insulators that were obviously built specifically for the rhombic station. Rhombic termination was accomplished with stainless steel wires strung up the middle of the rhombic. Other commercial stations have used "reentrant" termination, a term I have not yet found a definition for. Evidently reentrant termination permits the station to run unlimited power output without "frying" the termination medium (i.e. melting your stainless steel wire or your resistors)

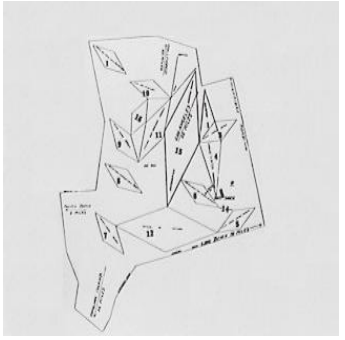
Click here to see the Rocky Point rhombic map with better resolution.



Amateur radio operators still occasionally employ rhombic antennas. In fact, I began my involvement after one day encountering the Don Wallace, W6AM, rhombic station. I at the time was a teenager in Palos Verdes, California. I was riding my bike down a country road and came across a huge installation of rhombics each perched upon telephone poles, some over 150 feet tall! (he used two 75 foot poles spliced butt to butt with redwood and angle iron splices to hold these monster poles together, he also did not like to waste pole length by burying pole in the ground and instead used a metal spike inserted into the bottom of the pole to secure the pole to the ground). The station in 1945 had 120 acres of land (when Don purchased it) with 16 rhombics, the longest being 1500 feet (end to end) and beaming Asia (his favorite antenna). This 1500 foot rhombic is widely believed to have been the largest beam antenna ever in amateur service. He had 61 telephone poles, each at 80 feet and another 90 feed line poles (25 feet high each). These antennas were fed by 52 MILES of feed lines (#8 copperweld). Several of these antennas had 2 or 3 curtains (to smooth impedance and increase gain). In 1962, Don sold 95 acres and consolidated his antennas. He installed ten 140 foot poles on the perimeter of his remaining 25 acres with rhombics layered every 6 feet in various directions with rhombics beginning at the 90 foot level and going in layers of 6 feet up to 140 feet high.

Click here to see the W6AM rhombic farm map image with better resolution.

Click here to read Don's handout he sent anyone inquiring of his rhombic farm

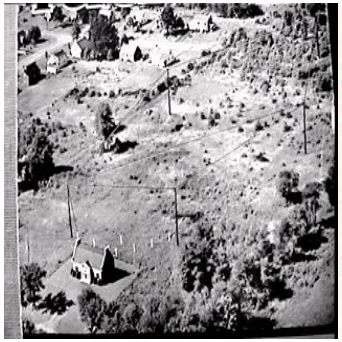


Now his longest (and still favorite) rhombic was 1100 feet end to end. Overall these antennas now covered 16 directions using the 8 rhombics he installed. This was the station I bicycled by. Don showed me his station with the miles of curtains and feedlines stretching for as far as the eye could see and with an operator's desk with a separate Collins station per band. He had his linear amplifier in a cage behind the operator's station and you would tune the amp with a fiberglass rod. The antenna feedlines crossed the ceiling of the station and when you changed directions, you would turn a heavy bakelite know on a rotary switch and the open frame relay matrix would chatter accordingly to your right. In the receiver you would hear various distant stations peak and fall as you electronically swept the horizon faster than any other operators could with their rotary arrays. Additionally, you would hear stations earlier during band openings and often you would communicate easily with far away stations others could not even hear. Nothing short of amazing. Don had an annual "open house" that was well attended. He always would fire up his spark gap transmitter especially for the occasion and demonstrate how a spark station sounded. I also recall Don's KW mobile station in his convertible Cadillac working Russian stations on the 495 Freeway at 60 MPH using an old bug strapped to the seat (on the way to Clifton's in downtown LA for the Southern California DX Association meeting). The code would sound slightly different as we would round steep turns

Eventually, after Don's passing, the station was dismantled for condomiums (Wallace Estates) and there is now a museum station there. At it's closure, the W6AM station had made over 500,000 contacts. I have a few of the W6AM insulators with which to remember this now silent station.

The American Radio Relay League (ARRL) in Newington Conecticut (not far from my home) had a rhombic until 1989, when dry rot in it's cedar support poles forced it to be dismantled (seems to be the fate of

most rhombics at some point, they outlive their supports!).



OUR COVER

We present an aerial eyeful of that ethereal eyeful, W1AW, ARRL headquarters station, shot by Staff Photographer Frank Benedin while flying at 600 feet. To the rear of the W1AW building are the five 65-foot antenna poles and the feed-line supports. The station occupies a 7-acre site at Newington, Conn., 4 miles from Hq.

Diagram of W1AW antennas: AB, 14 Mc. DE, 7 Mc.; AF, 28 Mc.; ABCDA, rhombic used mainly on 14 Mc. A 50-Mc. horizontal "Q" is mounted atop Pole E.

QST for

The four 65-foot red cedar poles that supported the famous W1AW Rhombic were taken down in August. The poles came from the West Coast and were transported through the Panama Canal. Installation took place in 1937 by a volunteer crew from the Hartford Electric Light Co. The poles had been the victims of noticeable dry rot. Here, a tree service employee cuts one of the antenna supports into sections before it's transported off the HQ site.

Click here to see the QST cover rhombic image with better resolution.

Click here to see the QST text description of the ARRL rhombic with better resolution.

Click here to see the image of the dismantling of the ARRL rhombic with better resolution.

This antenna was supported at 65 feet and had leg lengths of 350 feet (theoretical gain of 17dbd on 14 Mhz) with a tilt angle of 70 degrees. It was oriented East-West and is said to have had a stomping signal (loudest in the US as the one wavelength height limited the radiation angle to 15 degrees, about right to be optimum for the 1500 mile hop to the West Coast).

More recently, during the Desert Storm engagement, a group in the Midwest set up a rhombic beaming the Saudi Peninsula to provide point to point communications for the Military Amateur Radio Service (MARS). Again, the rhombic was chosen because of it's excellent point to point communications ability. This rhombic was up around 100 feet high with leg lengths of 294 feet (theoretic gain 15 dbd) and a tilt angle of 70 degrees. This antenna was said to permit eavesdropping on "manpack" tranceivers on the ground with rudimentary whip antennas (I don't doubt this).

Lastly, my modest rhombics. I have a "baby" rhombic that is 377 feet long (end to end) with leg lengths of 210 feet beaming East-West. The tilt angle is 64 degrees and because it is only 40 feet high ("on the ground" in rhombic terms), the radiation angle is around 25 to 30 degrees (very loud to the West coasts of the United States and Africa to say the least!). I also have a rhombic on an unoccupied property that is a bit bigger: 288 feet leg length, 520 feet end to end with a tilt angle of 65 degrees. It is also a bit low at 45 feet (tree supports). Someday I aspire to install a few 90 foot telephone poles but am still trying to figure out an economical way of getting/installing same (any help from out there in internet land?).

As time permits I will be scanning some rhombic-related images in to place here on my page. Some links:

Rhombic Antenna Design Software. This software is included here courtesy of Orrin Winton WN1Z. This program is a windows app and allows you to optimize rhombics for desired wave angles and gain. You will need VBRUN300.dll to install this program.

Visit Joby Wieser's Rhombic Ranch. Joby has a number of rhombic antennas running

Accumulated email related to rhombics and rhombic history

Plans for VHF and UHF Laporte rhombics

Read the original foreword of Harper's Rhombic Antenna Design textbook

Click here to read Harper's discussion of phasing pairs of rhombics

Click here to read Harper's discussion of lightning supression for rhombic antennas.

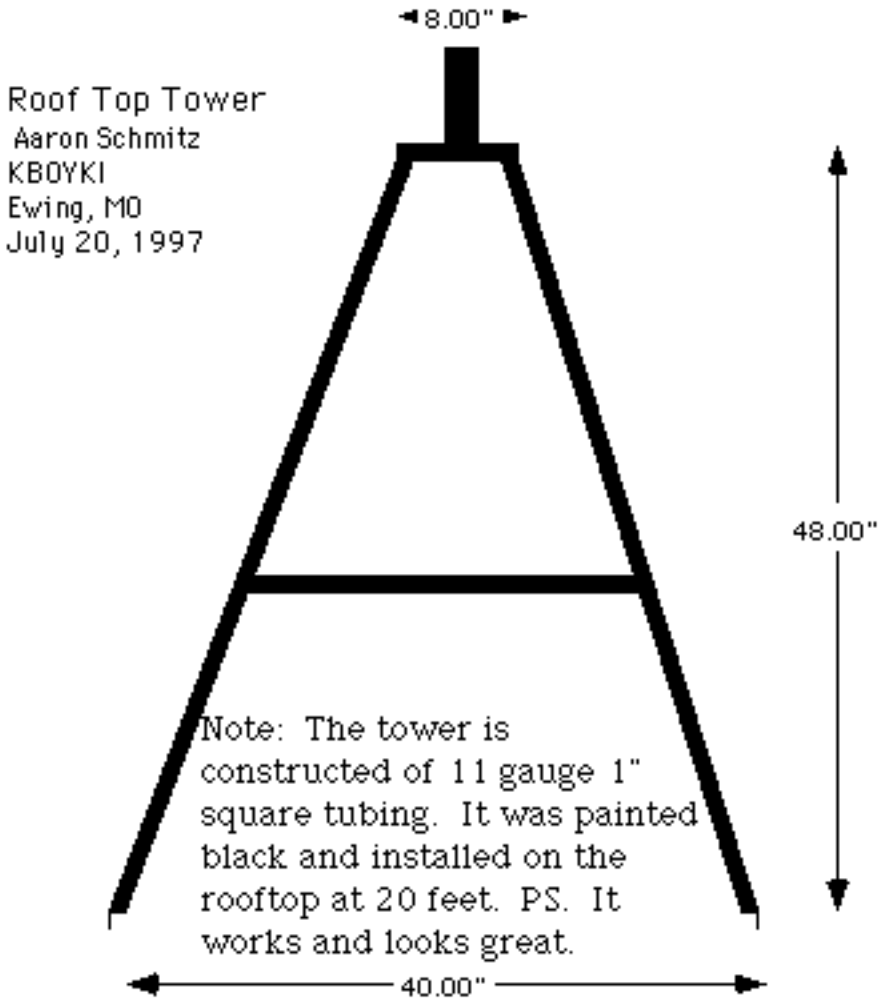
View the antenna patterns of various rhombic designs at various frequencies. I think these patterns bring a bit more clarity to how truely great the rhombic antenna really is.

American Radio Relay League (ARRL)The League is an organization comprised of radio amateurs with diverse interests. They publish QST magazine. I have been an ARRL member for over 25 years.

Email me at: cummings7@mindspring.com

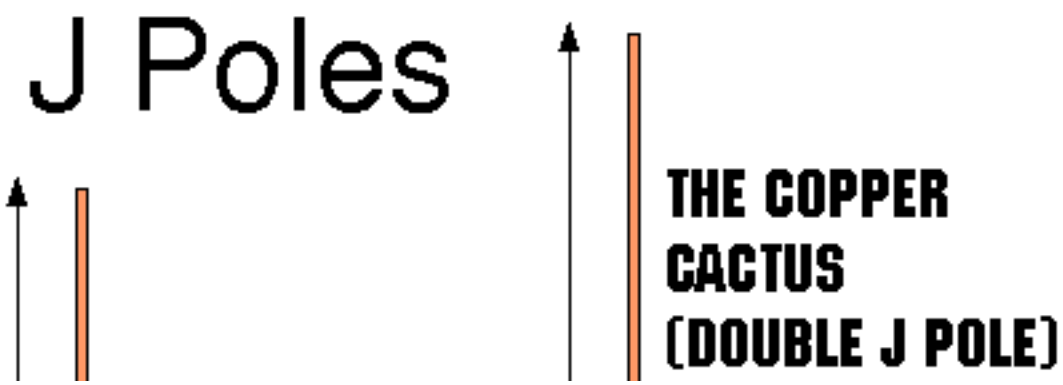
Return to my home page.

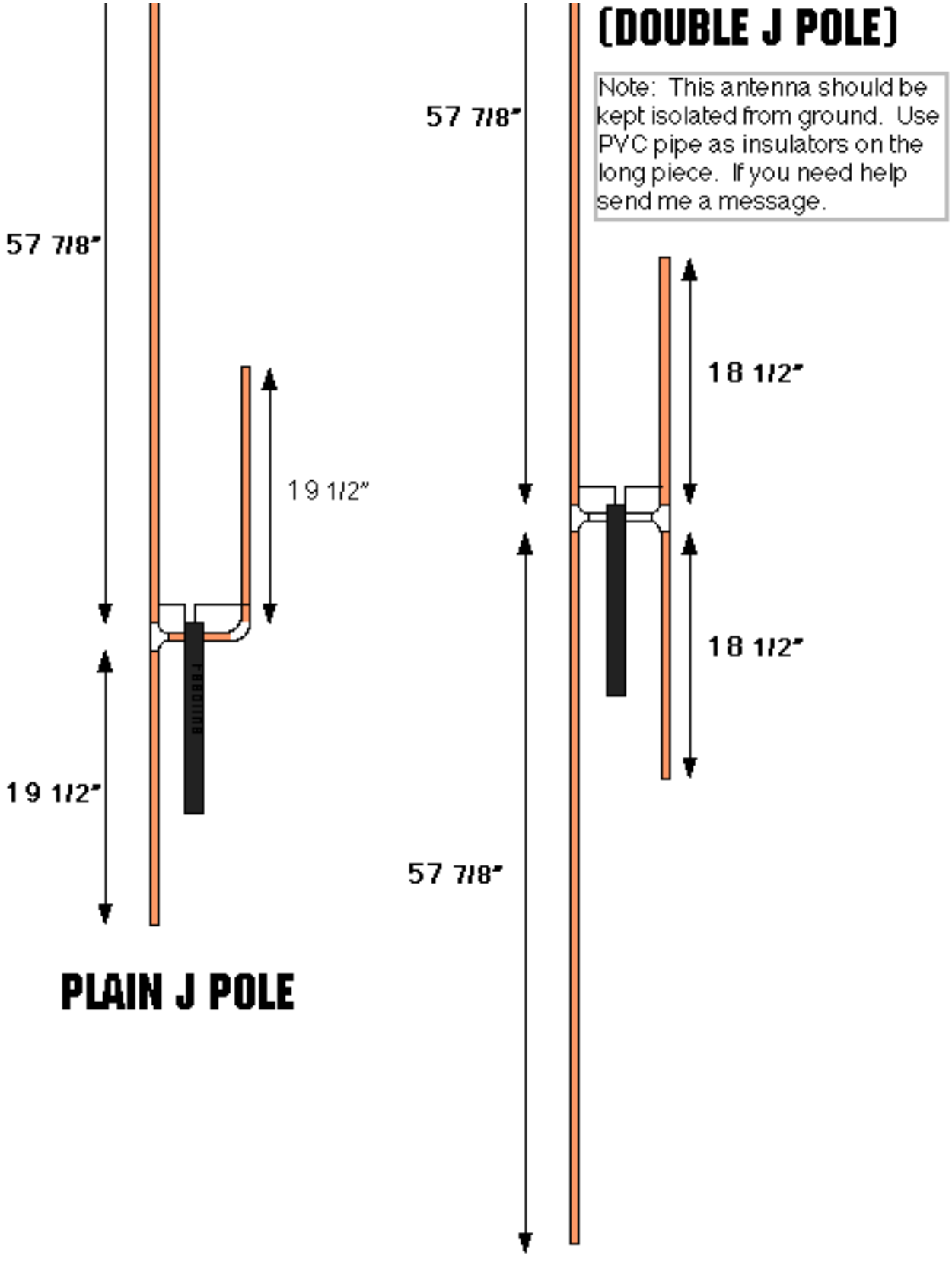
The Rooftop Tower



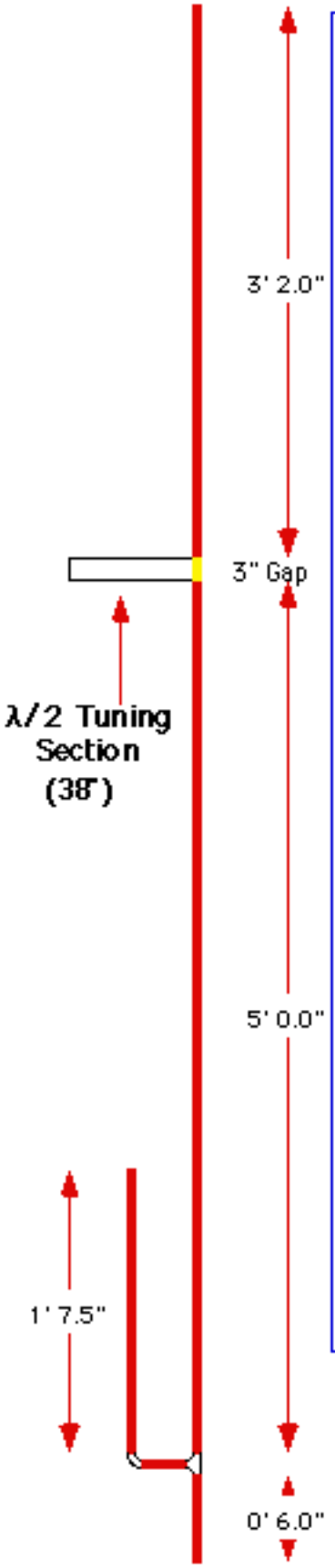
Common Amateur Radio Antennas

The J Pole





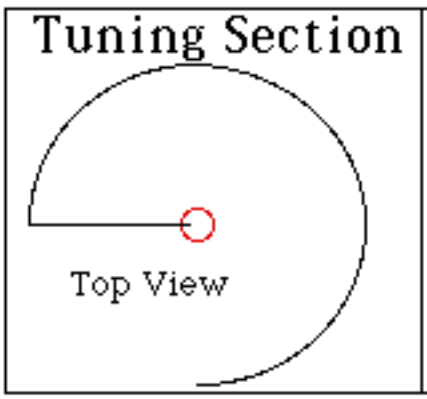
The Super J Pole

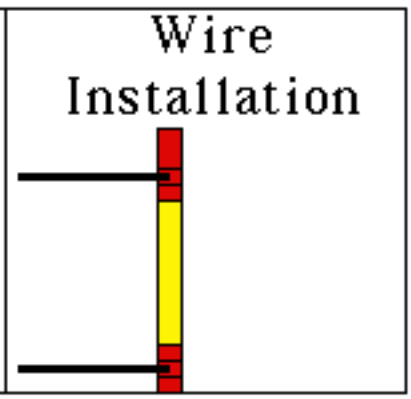


Super 2 Meter J Pole Antenna (a.k.a. Collinear Antenna) with helpful modifications by Aaron Schmitz, KB0YKI

Construction Details:

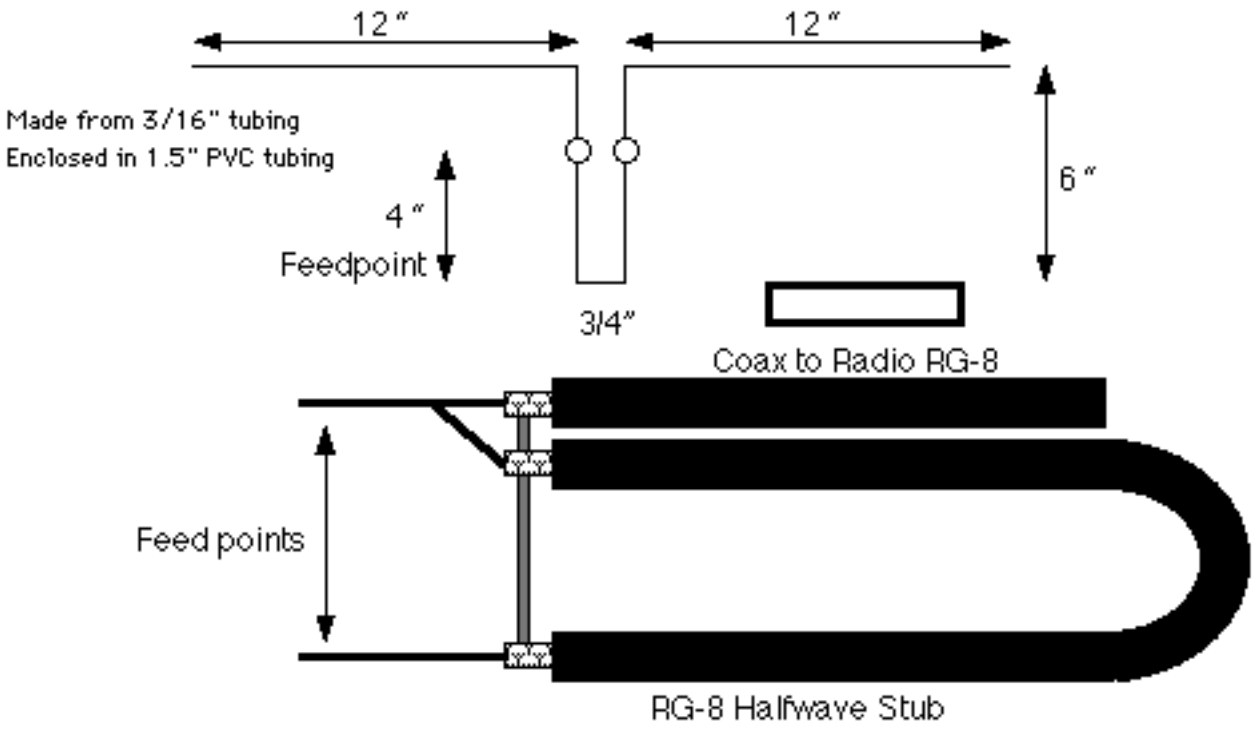
1. Cut copper to proper lengths.
2. Flux and solder bottom J Pole section.
3. Cut a 38" piece of 12 gauge wire.
4. Solder a 1/2" coupler on the top J pole section and bottom of the top λ/2 length of pipe.
5. Drill a 1/8" hole into each coupler through the pipe.
6. Strip back 3/8" of insulation on wire.
7. Insert wire ends into holes and solder.
8. Bend wire into a 3/4 circle around pipe with a 6" radius.
9. Insert 1/2" X 4" dowel rod into coupler.
10. Completely seal all areas from moisture. Paint it if you would like.
11. Attach coax and have fun!!





Omnidirectional

440-450 MHZ Omnidirectional Antenna
from
73 Magazine November 1994

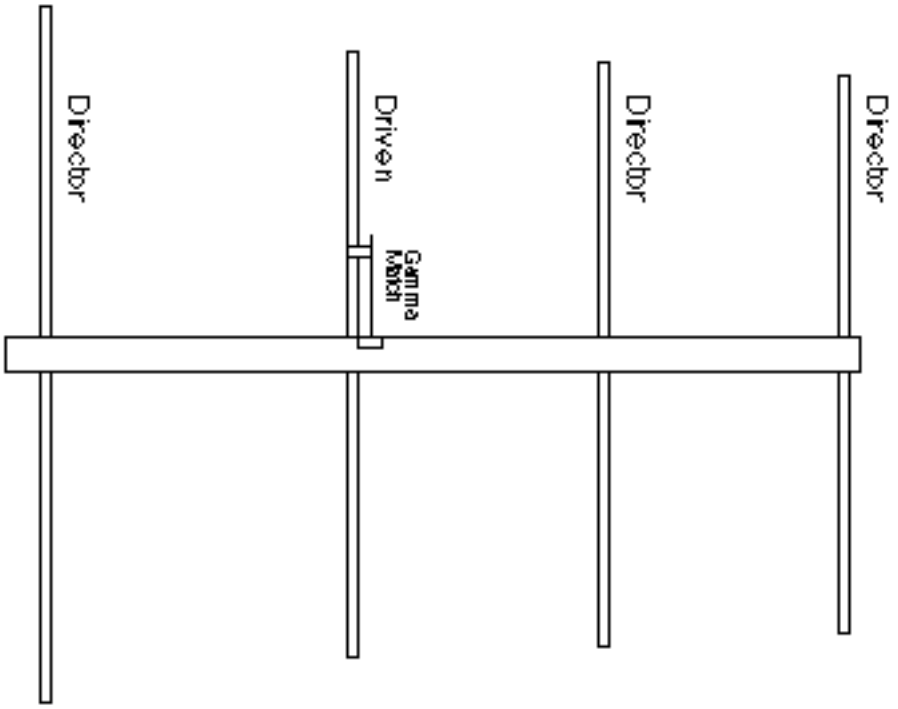


Yagi

Element	Length	Spacing
Reflector	19.78*2	
Driven	19.25*2	15 5/16"
Director 1	18.25*2	15 5/16"
Director 2	17.125*2	15 5/16"

Antenna Made from arrow shafts

Yagi
April 1992 edition of 73 Magazine



Quads

VHF AND UHF QUAD ANTENNAS

Plumber's Quad (4 Element

	146.000 Element Length (cm)	Element Spacing (cm)	146.000 Element Length (in)	Spacing (in)	446.00 MHz Element Length (in)	Element Spacing (in)
Reflector	216.000	40.000	85.039	15.748	27.983	5.182
Driven	206.800		81.417		26.791	
Director	200.700	39.600	79.016	15.591	26.001	5.130
Director	200.000	41.000	78.740	16.142	25.910	5.312

Quad Antenna 7 Element

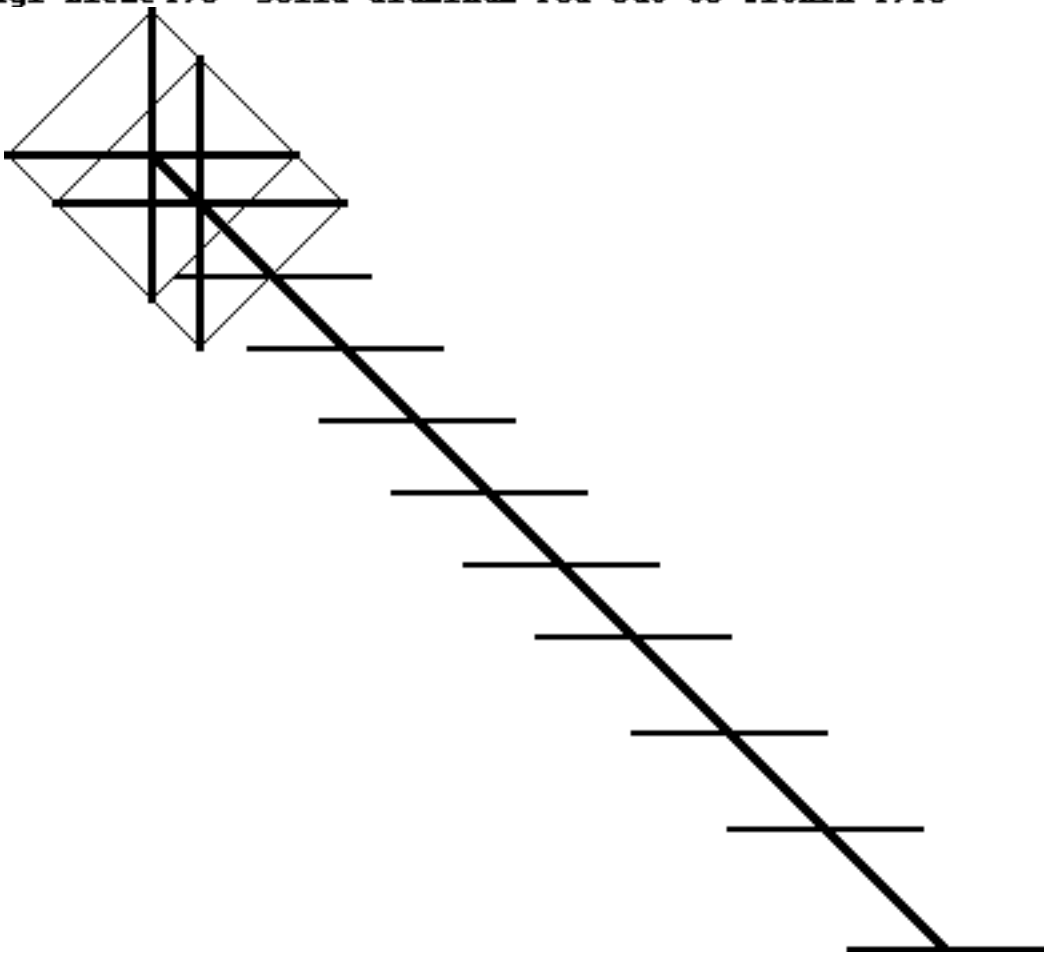
	Element Spacing 18"		Element Spacing 6 in	
	146.3 MHz Element Length (in)	Cross Piece (in)	446.00 MHz Element Length (in)	Cross Piece (in)
Reflector	86.000	30.000	28.206	9.839
Driven	82.000	28.500	26.894	9.347
Director	78.000	27.125	25.582	8.896
Director	77.500	27.000	25.418	8.855
Director	77.000	26.750	25.254	8.773
Director	76.500	26.625	25.090	8.732
Director	76.000	26.500	24.926	8.691

Quagi

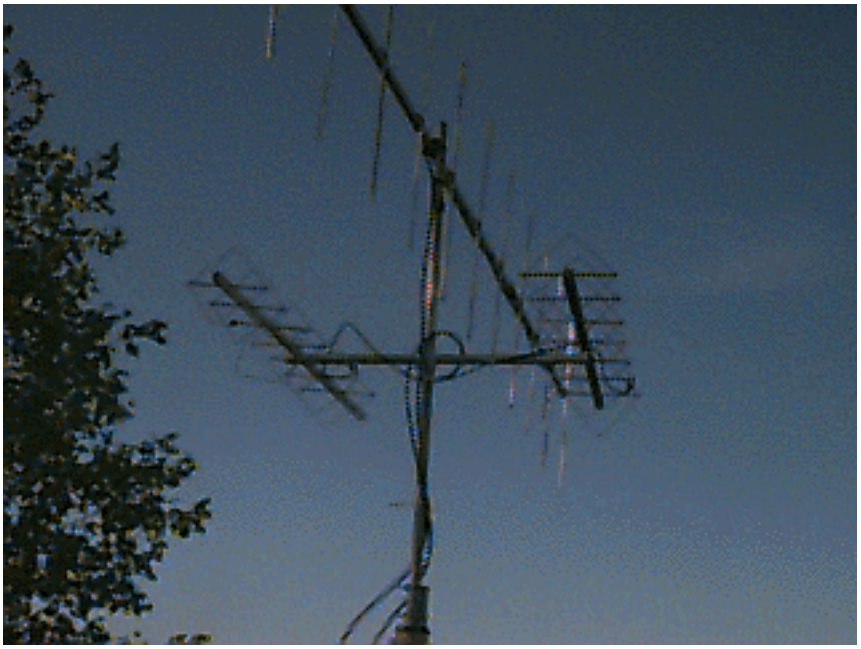
Modified W5UN Quagi Design
Optimized for 144.050 MHz

	Element Le (inch)	Spacing to element (inch)
Reflector	86.7500	
Driven Ele	82.0000	21.0000
Director 1	35.9375	15.5000
Director 2	35.7500	33.0000
Director 3	35.3750	31.0000
Director 4	35.2500	33.0000
Director 5	35.0000	31.6250
Director 6	34.8125	36.0000
Director 7	34.6250	36.0000
Director 8	34.6250	34.0000
Director 9	34.5000	34.5000

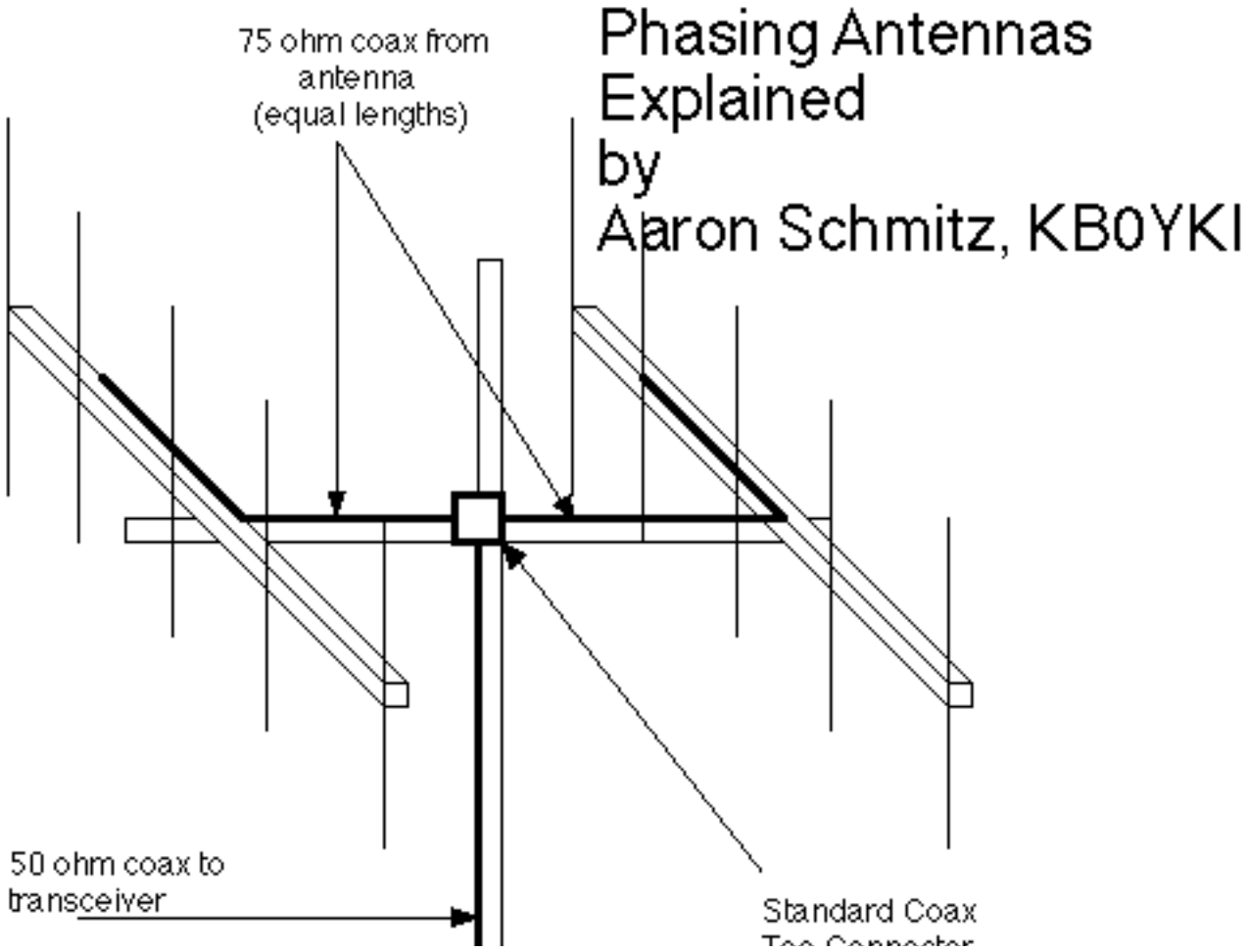
Boom- Nonconductive material such as wood or fiberglass
Quad Eleme made from 12 guage insulated copper wire
Yagi Eleme 1/8" solid aluminum rod cut to within 1/16"



My Stacked Quads



The picture above is of my 440 MHz stacked quads. The spacing between the antennas is approximately 80-90% of the boom length. RG-59 (75 ohm) cable was used on the phasing harness. An almost perfect 1:1 SWR was achieved at 446 MHz after installation. Any other questions on this can be forwarded to me.



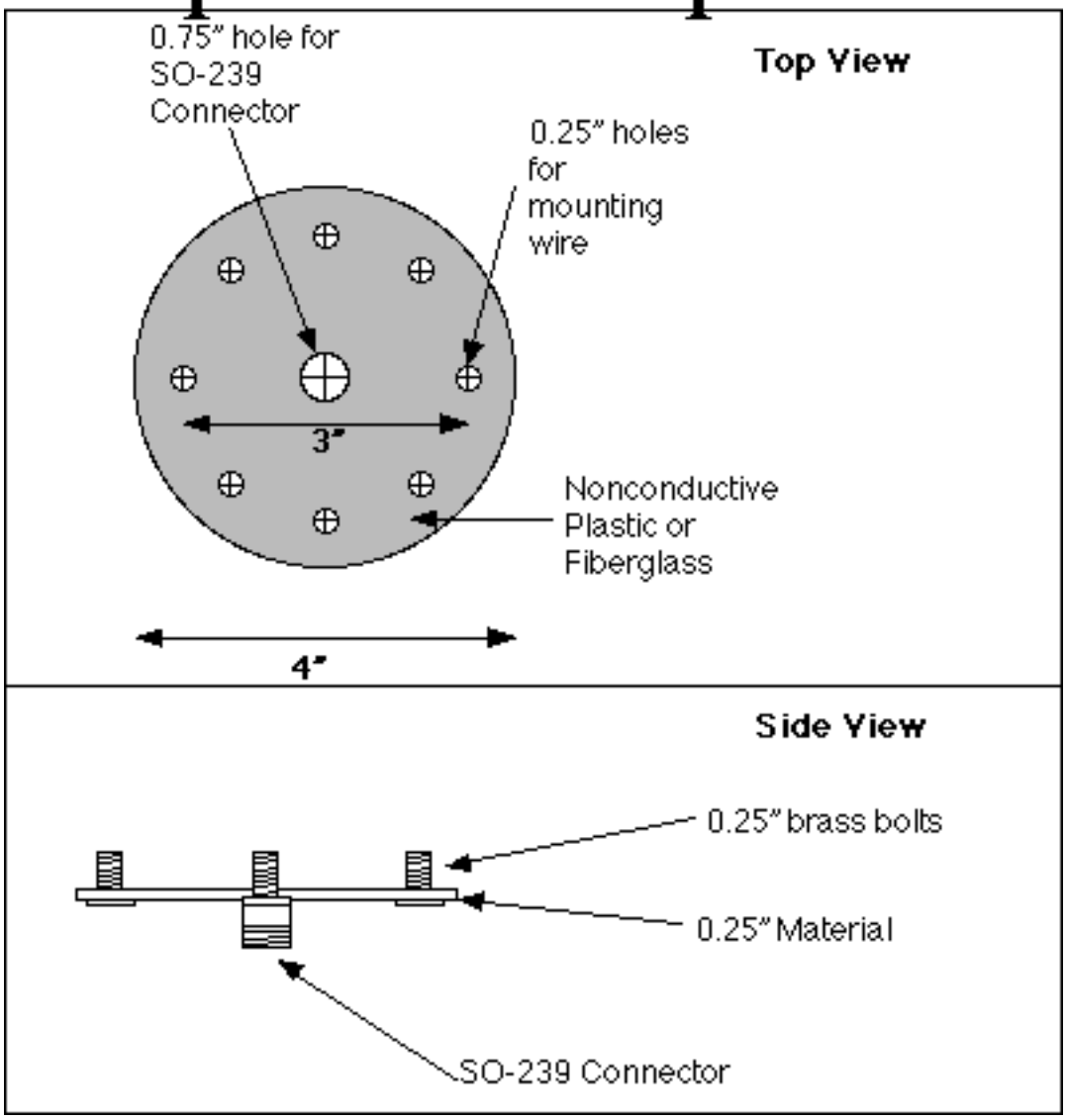
Standard Coax
Tee Connector

Steps to build a Phasing Harness
1) Build the antenna and cross support
2) Determine the length of coax that you need to run from the antenna feedpoint to the connector.
3) See how much coax will work for the phasing harness.
f=146.520 MHz
λ/4=234/f=234/146.52=1.5 ft=18 inches (Physical λ/4)
RG-59 Coax Velocity factor is 66% the speed of light.
(18 inches)X(0.66)=12 inches (Electrical λ/4)

Example: Your setup needs approximately 6 feet of cable to get from the feedpoint to the Tee connector.
The only problem is that 6 ft would be an even quarter wavelength (not good). All you have to do is go to 7 feet of RG-59 coax and you have 7 electrical quarter wavelengths(Fantastic). You should be pretty close to having a low SWR reading. That is not guaranteed, but it should be close.

4) Get your 75 ohm coax and cut it off pretty close to amount of coax needed(in this case 7 feet for each side.)
5) Solder on your PL-259 connectors with the appropriate reducer to the 75 ohm coax coax.
6) Hook everything up. If you are using a gamma matched antenna, adjust the gammas to get a low SWR reading.
7) Get on the air and have fun!!!

Dipole Feedpoint



This is the big rooftop tower that I built a few months ago. At that time I could not find any in my area for a price I felt I could afford. It is made of 1 inch square tubing for the legs and uses a 2 1/4" inch pipe for holding the rotor. It is welded throughout and has two coats of enamel primer and two coats of flat black paint.

The tower has with stood the 50 mile per hour winds that have roared through Missouri in the month of April and May of 1997. If you would like a set of plans for this, feel free to email me.

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A ROTARY DIPOLE FOR 17 AND 20

After assembled a two element Quad, Mario (ik7zcq) needed an antenna for 17 and 20, so he asked me to build a 17/20 dipole; I hote rotary dipoles and traps..but why don't try it!

THE PROJECT

I hote projects because I never find all I need: so always I do the best I can with what I found.
At the beggining my project was:
1) Grounded-driven element;
2) Gamma-match feeding;
3) Traps with coax-cable;
After first test I decided to use a direct-feeding with no Gamma-match: the Gamma will give best results with single-band feeding. Dipole will be no more grounded.

THE DRIVEN ELEMENT

Driven element is tapered and I use 8 aluminum pipes with the following diameters:

Diameter	Length	thickness	N.part
30 mm	162 cm	2 mm	2
25 mm	164 cm	2 mm	2
20 mm	131 cm *	2 mm	2
16 mm	28 cm	1 mm	2

*** here will be the trap (108 cm + insulator + 23 cm)**

I suggest you to reduce thickness for tips (2 - 1,75 - 1,5 - 1 mm).

First part (diam. 30 mm) is center-insulated with a nylon bar (diam. 25 mm);

outside I used PVC (diam. 40mm, 5 mm Thickness); there are two screw bolt (3mm diam.) for feeding the dipole (See Picture 1).
The mast to boom attachment is an home-made steel plate (15x30x0,35 cm) (See Picture 2).

Second part (diam. 25 mm) is easy to assemble;

Third one is a little bit difficult: cut a 20 mm aluminum pipe (lenght 131 cm) at 108 cm so to obtain two pipes (108 e 23 cm).
Put a PVC bar (5cm) into and between them, then use fibreglass to block them and put over them an insulated support (See Picture 3) (TNX Antonio,ik7ytx).
Than make two 3 mm-holes (for traps-feeding).

Last part is a 28 cm length (diam. 16 mm) aluminum pipe, used for 20 mt tuning.

TUNING

Radiation resistance for an half-wave dipole will be around 69-70 ohm (free space); if you put it 0,5/0,75 lambda high resistance will be lower so you don't need to match it; I only used 20-25 turns of coax-cable as balun.

THE TRAPS

At the beginning I believed traps was the more difficult thing to build, so i bought two RF capacitor (See Picture 4) but that was not true!
Using a dip-meter check the resonance: 8 turns of RG58 on a 27 mm support will resonate at 18.1 (See Picture 5).
When tuned, cover the traps with RF glue (See Picture 6) so that water can't enter in and then cover all with PVC tape (See Picture 7).

ASSEMBLING

I tested the dipole on my garden, on a 10 mt pole, with no objectes near it.
At first assemble first two partes (with no traps) and tune it on 17 mt; than add traps and 20 mt part and tune it on 20: S.W.R. will not change.
On my garden S.W.R was 1,2:1 on 17 and 1,5:1 on 14.200, 2:1 on 14.000 and

14.350; when assembled in the qth of Mario it was 2:1 on 17 and 1,1:1 on 14.200 (but there are some objects near it, see Picture 8)!

The length (half-dipole) is the following:

Tips (18 MHZ)	Trap	Half-Dipole (18 e 14 MHZ)
46 cm	Trap	394 cm

WEIGTH

Dipole's weigth is less than 6 Kg but you can reduce it reducing the thickness for aluminum pipes.

73 de iz7ath, Talino

RUBBER DUCK ANTENNAS

RACES BULLETINS 023-025, 1986

Rubber duck antennas on hand-held radios are a severe compromise on efficiency. On the plus side is their short size and flexible forgiveness to brutish handling. On the negative side is their terrible radiation inefficiency, probably worse than many of you expected.

When did you last replace your helical spring antenna we call the rubber duck? On testing a hundred or so portable radios that had been out on the fire lines for a few weeks we found a typical 60 percent failure rate. Most of the antennas looked fine. The only way you can detect an invisible rubber duck failure is by measuring the microvolts per meter with a calibrated receiver over a measured range under controlled conditions, such as done routinely by the Boise Interagency Fire Center. Since this is difficult for most to do, it might not be a bad idea to replace rubber ducks as a matter or course when they show signs of wear or if they are a year old. You might want to consider using a telescopic antenna under non-violent conditions to vastly improve the range of your hand-held.

The National Bureau of Standards ran some tests that proved what we had long suspected. The efficiency of a hand-held is dependent upon how much antenna it has and how good the ground plane. Most portables have very poor ground planes; the more metal the better. Also the more antenna the better. Hence the rubber duck is a woeful but often necessary compromise. But if a portable is not going to be subjected to the abuse of fireground or street cop utility, you should consider the telescopic quarter-wave antenna if range is important. Compare the figures and discussion that follows.

Be aware that the telescopic antenna is nowhere as rugged as the rubber duck but it will talk circles around it. You might say that the quarter wave whip is to the rubber duck what a 106 inch CB quarter wave whip is to a 36 inch whip on a base loaded coil to compromise range for low garages. Our reference antenna in the Public Safety high band and 2-meter Amateur radio measurements below is a quarter-wave telescopic antenna, extended, and held at face level: One-quarter wavelength extended and at face level = 0 dB One-quarter wavelength collapsed and worn at belt level = -40dB Rubber duck held at face level = -5dB Rubber duck worn at belt level = -20dB Translated, this means that a 5-watt hand-held with a rubber duck worn on the belt has an effective radiated power not of 5 watts but only .05 watt. Held at face level the radio has an ERP of 1.6 watt. 15dB is quite a difference!

In the material above we gave you facts and figures of the quarter-wave telescopic versus the rubber duck for Public Safety VHF Highband and 2-Meter Amateur handhelds. The 40 dB down for the nested telescopic relates to those commercial models where the telescopic disappears within the radio. Such an antenna won't break when it's nested but it won't receive worth a whoop either. In those radios where the collapsed quarter wave is external to the radio they break very easily. For that reason we recommend the style that has a spring at the base. The spring makes it very forgiving of elbows and other bum raps. We

have not researched or measured five-eighth wave antennas because they are too long for most public safety use and because they typically require too many telescopic sections. The more sections the more chance of troubles. Few people take the time to correctly telescope any hand-held antenna. They should never be whacked down with the palm of the hand on top and push. They should be pulled down with the thumb and first two fingers.

If you are interested in the figures for 450 MHz, using the table above, they are respectively 0dB, 30dB, 5dB, and 30dB. One more reference for the technically inclined-the loss of a telescopic antenna compared to half-wave dipole: VHF -5dBd and UHF -20 dBd. Telescopic antennas should be changed at least annually and whenever they become the slightest bit loose. Any looseness can mean a poor RF connection inside the antenna where you can't see it or fix it. Simply change it.

Archives of California RACES Bulletins are available via anonymous ftp at ftp.ucsd.edu/hamradio/races

The Home Page for the State of California Governor's Office of Emergency Services contains other information about emergency response in California and elsewhere.

Use your Back key to return to previous pages

The Slinky that became a Antenna

LOWER PRICE WAS $35.00 *NOW* $29.95

Fun with Helitrix ™, the stretchy helix.-- Helitrix ™ is a flexible springy helix with a 2-3/4 inch diameter. Each Helitrix ™ weighs approximately 7 ounces. When collapsed it is only 2-1/4 inch long, but it can be stretched without deforming into a helix as long as 15 feet in length. An antenna made from a Helitrix™ is light, simple to suspend and extend, and easy to put out of sight when not in use.

Helitrix™ Antenna Basics.-- For a forty meter dipole, use a full Helitrix™ coil on each side of the feedpoint. Stretch to 15 ft. For 75 meters put two coils on each side and stretch the whole to about 30 feet in length. Presto! You will have an 80 meter dipole that fits in most attics or even in a motel room. A quick twenty meter dipole, can be made by stretching out a single coil to a length of 7 ft and feeding it with a delta match in the center.

Helitrix Antenna Performance.--Is Helitrix a wonder antenna? No. But it works. In a test on a state-wide 75 meter phone net, a 30 foot 80 meter Helitrix™ dipole at 20 ft received signal reports on average 1-1/2 S-units lower than a TNT Windom at 35 feet. That's not bad for a 1/8 wavelength antenna. The Helitrix dipole would have compared better had it been equally high, but it would not outperform the full size dipole. On the same occasion a comparison was made with a Hustler mobile whip. Compared with the whip, Helitrix performance and bandwidth were outstanding--better by several S-units for state-wide coverage. So, considering that you can even install a Helitrix ™ antenna for 80 meters inside a motel room, the Helitrix antenna promises some good moments.

1. The simplest way to obtain multi-band results is to take a pair of Helitrix™ coils, one on each side of your feedpoint, stretch them as far as space permits, and clip on the feedline. Here you have a compact version of the good old "center fed Zepp". Feed it through an antenna tuner. This simple antenna will play on all bands 7 MHz and above. In a pinch it will permit QSOs on the 80 meter band. The ready-to go QRV Helitrix 40-10 antenna (see description below) is an aesthetically pleasing and easy to use version of this basic doublet.
2. Helitrix™ antennas are best suited to indoor or portable deployment. If you wish to put your Helitrix ™ antenna outside on a more or less permanent basis, you should solder all connections well, then paint the whole antenna with spray enamel. This is actually quite easy to do, so don't shy away from trying it.
3. For a given Helitrix™ antenna, performance is best at the frequency of natural resonance and on the next harmonic because the coils act increasingly like an RF choke on the higher harmonics.

Have Fun with a Helitrix! But be careful the kids don't run off with it. They could slink around with it like a toy. But they'd miss the big fun. Could they imagine talking into one and having someone from Japan answer them on 15 meters? Or think of checking into MARS and RACES nets with a Helitrix™ on the bedroom wall? Only in amateur radio can you get a bang like this out of a stretchy helix.

Antennas & More has put together what you need to get off to a fast start with Helitrix™ experimentation. Check out the items listed below and turn a weekend with a Helitrix™ into the most fun you've had since you were a kid .

Helitrix™ Antenna TechNote 123D $7 ppd.
This frequently updated Technote reviews Helitrix™ experiments performed to date. Gives detailed information on the electrical properties of Helitrix™ coil antennas. Describes instruments used & designs tested. Reports

unpublished experiences of numerous contributors. Shows how to get the most out of a Helitrix™ with or without an antenna tuner. Reviews what works and what doesn't. Use this to study out the principles and plan your installation.

QRV Helitrix™ Dipole $29.95

This is the fully refined and ready to use QRV Helitrix™ system, all set to go for instant motel, attic, or portable use. Coaxial feedline is decorator white. Quick attachment clips make it a snap to hook up fast. Support line lets antenna slide in and out of action instantaneously. Unique design assures unerring and quick return to correct position whenever desired. Mounting tabs, coax feedline, and unobtrusive installation easy, whether for long or short term utilization. Manual shows proven methods for working 40 thru 10 even without an antenna tuner. With tuner work 80 meters.

Includes Feedline

Helitrix 40-10M Dipole antenna $29.95us

Helitrix 80M Extension $8.95us

Review Cart

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Send mail to webmaster@antennasmore.com with questions or comments about this web site.

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Last modified: April 11, 2004

AMERICAN QRP CLUB

Projects

Some tools, some rigs, some instructional, some published ... all interesting!

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NorCal Projects

CW Decoder Program for Your PC
Full featured CW decoder program for your computer, by Grant Connell, WD6CNF ***Download file updated! (3/9/04)***

Crappie Antenna
Use this collapsible fiberglass fishing pole as a neat antenna support, as Howard Zehr N9AHQ describes.

Antennas
Some rules of thumb for beginners, described by James Duffy, KK6MC/5.

Software
This is the (updated) set of BASIC programs mentioned in the article "RF Fun with the RF-1" published in the Spring 97 issue of QRPp.

Toroid Design Tool
... by Jack Ponton, GM0RWU

NB610
A homebrew 10m transceiver, by Wayne McFee, NB6M

SMK-1 on 20 Meters
Taking the venerable SMK-1 design to 20m and adding an RF amp, audio filter, and a TiCK Keyer ... by Wayne McFee, NB6M

Solid State VXO, Buffer Amplifier, and Keying Circuit For Tube Rigs
Useful accessory allow the use of smaller, modern crystal types to control and provide drive for tube rigs ... by Wayne McFee, NB6M

NJQRP Projects

Digital QRP Homebrewing
Home site for the article series on the Digital QRP Breadboard, the HC908 Daughtercard and other featured digital projects.

Back to the Future: Tuna Tin 2 & Herring Aid 5
Vintage QRP transmitter & receiver projects resurrected & updated by NorCal

Build the KA5DVS "PAC-12" Antenna
A homebrew, multiband portable vertical antenna.

Build a Logging Frequency Counter
Turn the PSK31 Audio Beacon project into an SX28-based frequency counter that talks to your PC

Antenna Analyzer I
Here's the project that started N2APB & N2CX on the journey: an inexpensive scanning VCO and SWR bridge

Simple Regen Receiver
The N1TEV design from QST was built up by a rural Maryland Boy Scout troop, with great results.

Portable PSK
N2APB's quest for a portable, stand-alone controller that does PSK31 encoding/decoding on 80 & 40 without using a PC

Summary of Sierra Mods
A collection of modifications that can be made to the ever-popular Sierra QRP Transceiver

How Low Can We Go, With a VXO

I wanted to learn more about how a VXO with a wide tuning range could be used in either Direct Conversion or Superhet transceiver schemes.... by Wayne McFee, NB6M

NB6M Homebrew DT Filter es TX

... by Wayne McFee, NB6M

NB6M Homebrew Receiver and Keying

... by Wayne McFee, NB6M

NB6M Homebrew VXO

... by Wayne McFee, NB6M

St. Louis Vertical

... by Dave Gauding, NF0R

St. Louis Vertical Updates

... Compiled by Jerry Parker, WA6OWR

NB6M Paddles

... by Wayne McFee, NB6M

Pixie-II

Collection of useful information for homebrewing this super simple QRPp transceiver

MicroBeacon

Collaborative project by NJQRP members featured a microcontrolled memory keyer with a controllable RF attentuator.

Homebrew Kites

This homebrew Scott Sled design from AA1MY takes HF antennas high into the air

DDS VFO with AD9850 + PIC16C84

N2APB homebrewed the classic DDS VFO project from WV2B published in July 1997 issue of QEX

Cigarette Lighter RF Probe

A QRP accessory for under $10, by KW2R

NN1G Transceiver Construction Notes

KW2R describes his notes on this classic transceiver from Dave Benson.

Antenna Musings

Excerpts from a presentation N2CX gave at FDIM 97 describing portable antennas

QRPpaddles

WK8G details his marvelous little paddle design

NorCal / K8FF Paddles

Some notes and photos from N2WF concerning his construction of the NorCal Paddles Kit

Z-Match

Dave Maliniak, AA2A collects his notes and the original article on the venerable ATU design from W6JJZ

The Perfect QRP Enclosures

N2APB found a commercial enclosure that was perfect for many QRP station accessories

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These pages are designed and maintained by George Heron, N2APB
Page last updated: Jan 4, 2004

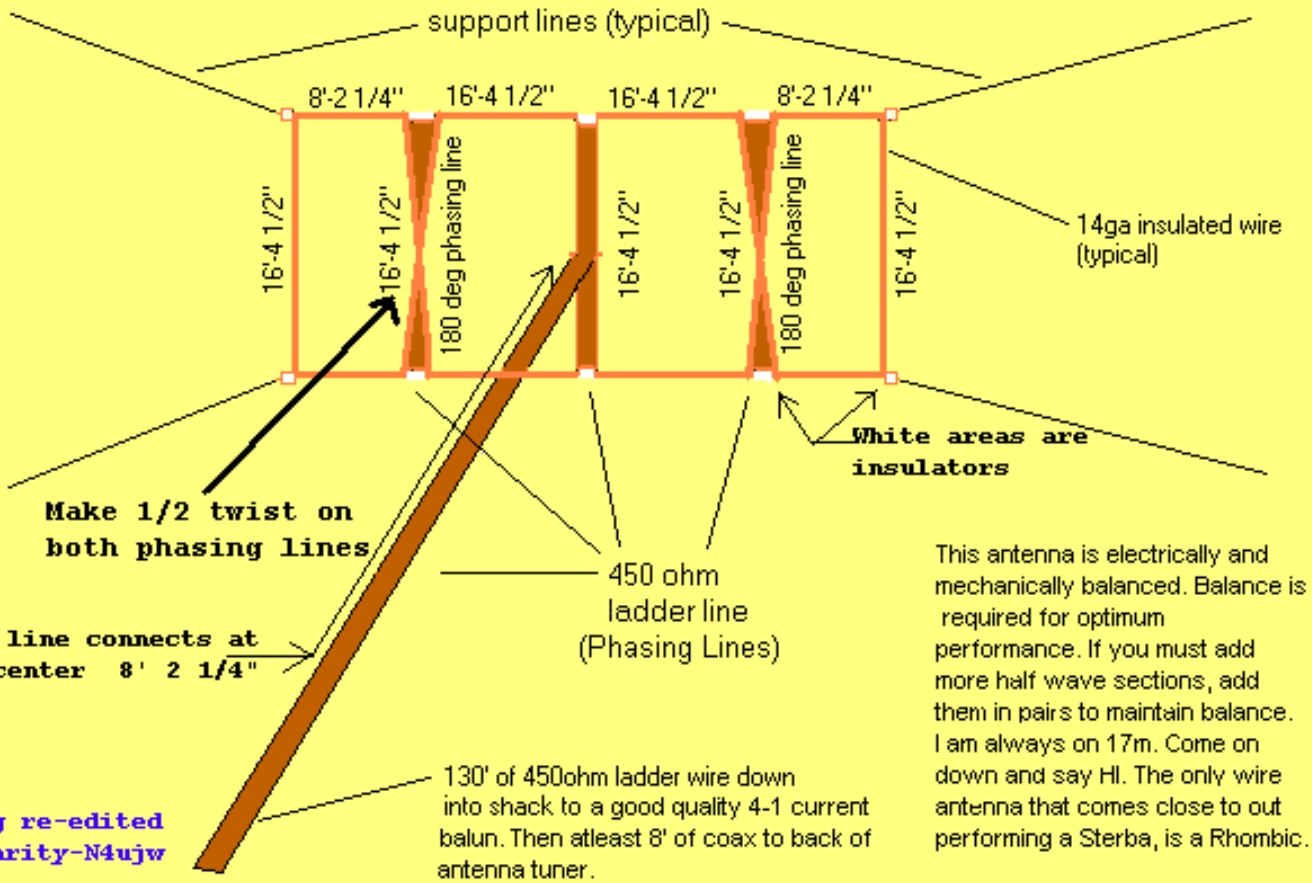
160m-10m Sterba Curtain

IF YOU COULD ONLY PUT UP ONE ANTENNA...THIS WOULD BE IT!

By NØKHQ - ST.LOUIS.MO

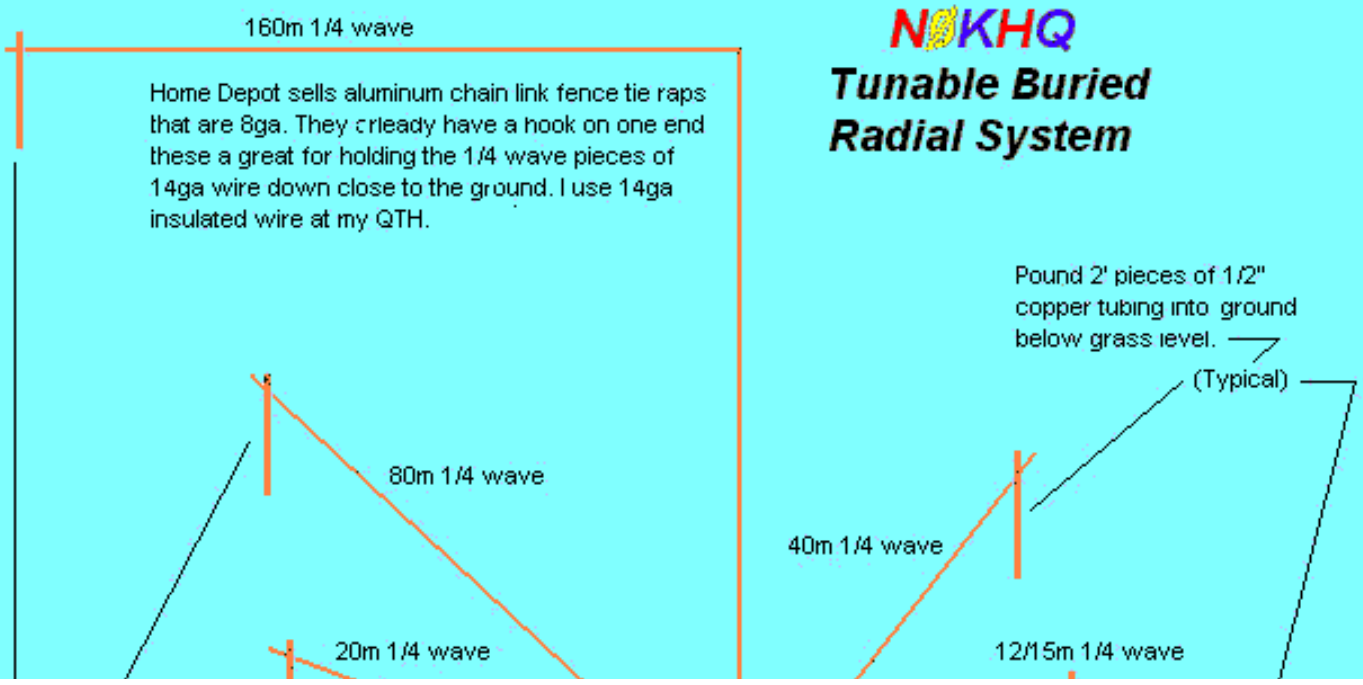
The Broadside Curtain Aray

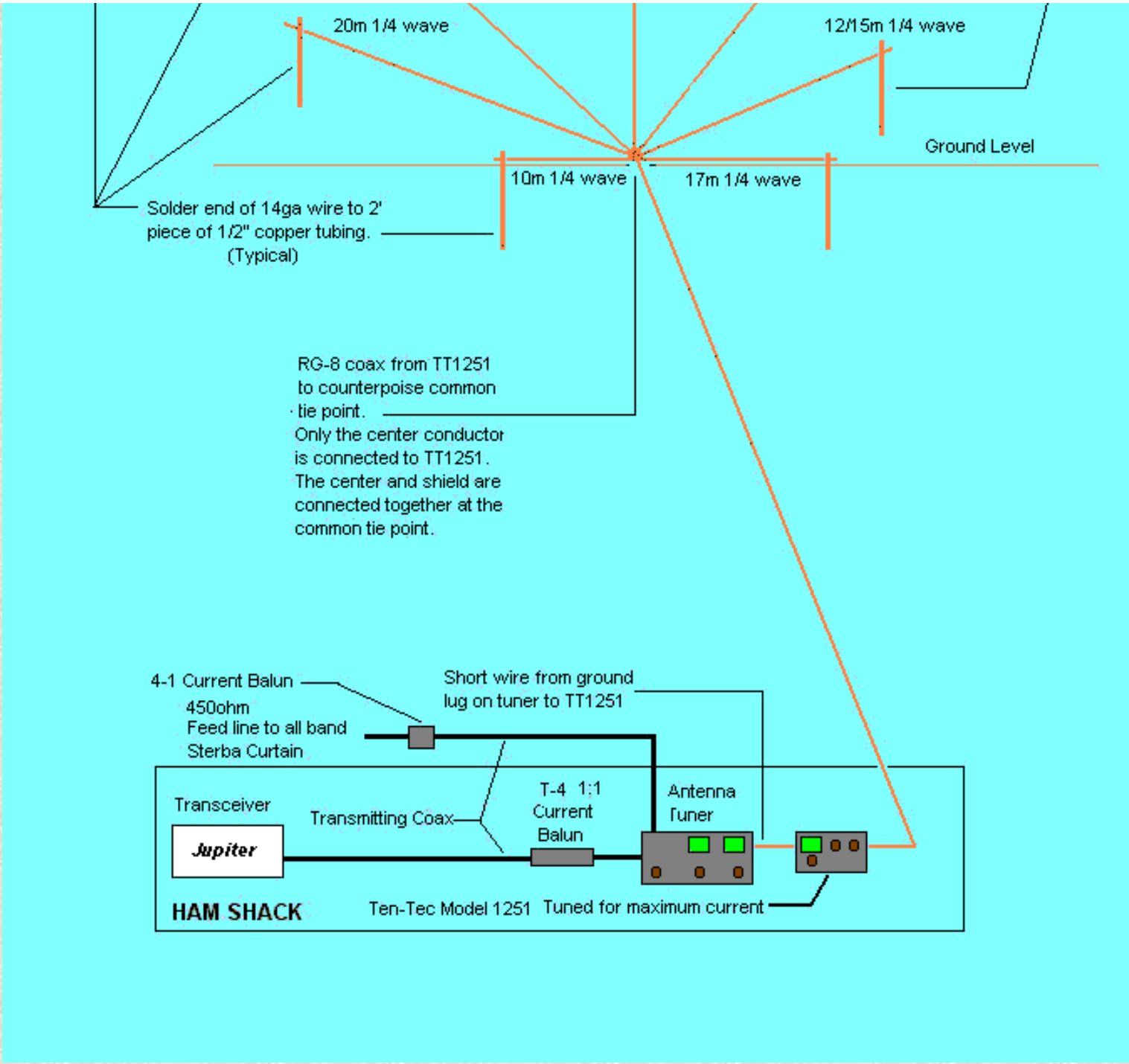
You can build them better than you can buy them!

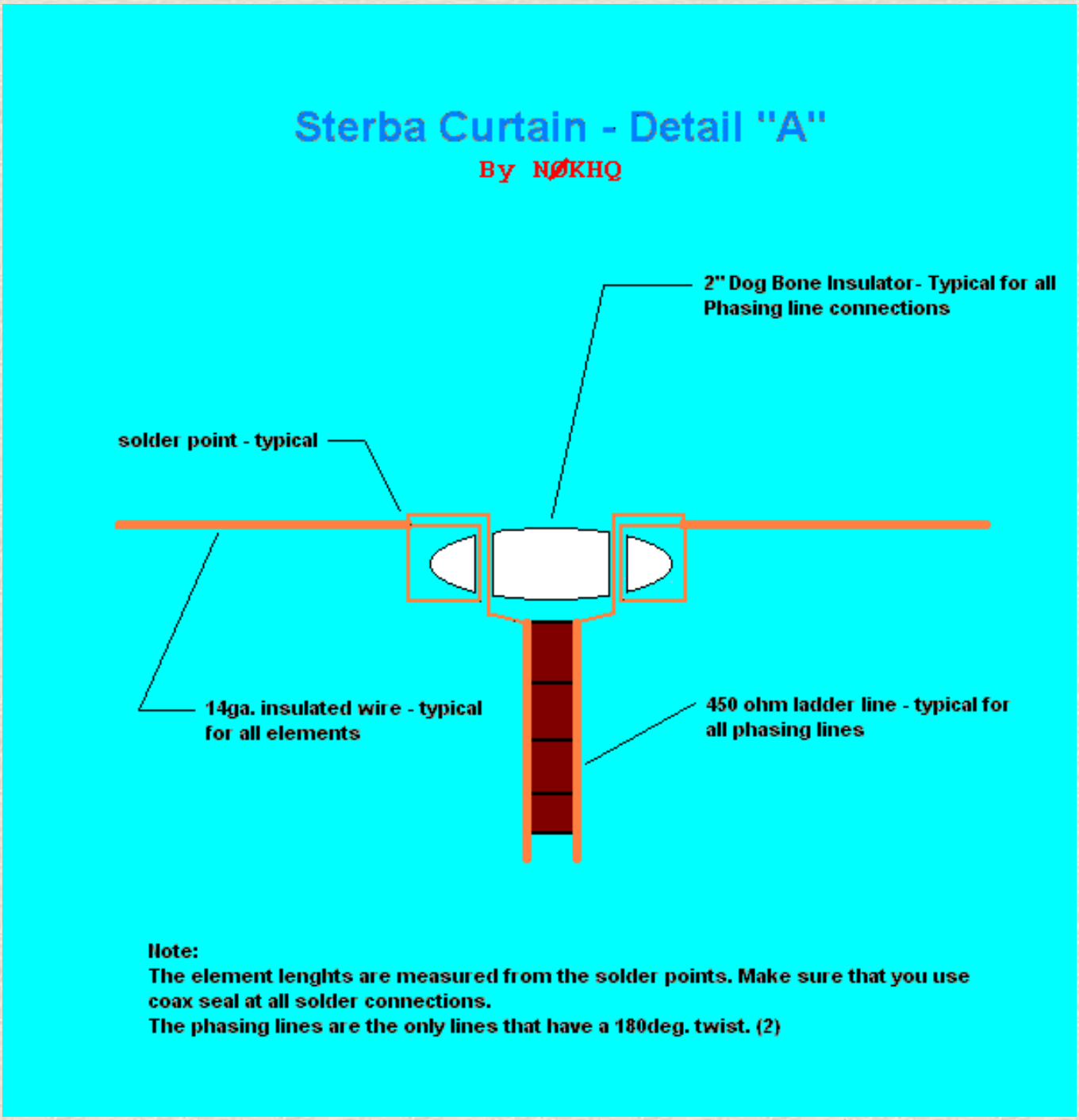


Drawing re-edited for clarity-N4ujw

This antenna is electrically and mechanically balanced. Balance is required for optimum performance. If you must add more half wave sections, add them in pairs to maintain balance. I am always on 17m. Come on down and say Hi. The only wire antenna that comes close to out performing a Sterba, is a Rhombic.




20m 1/4 wave
12/15m 1/4 wave
Ground Level
10m 1/4 wave
17m 1/4 wave
Solder end of 14ga wire to 2' piece of 1/2" copper tubing. (Typical)
RG-8 coax from TT1251 to counterpoise common tie point.
Only the center conductor is connected to TT1251. The center and shield are connected together at the common tie point.
4-1 Current Balun
450ohm
Feed line to all band Sterba Curtain
Short wire from ground lug on tuner to TT1251
Transceiver
Jupiter
Transmitting Coax
T-4 1:1 Current Balun
Antenna Tuner
HAM SHACK
Ten-Tec Model 1251
Tuned for maximum current



"I live on a very small residential lot 65' wide x 100' deep. I have a tree in the front yard and a tree in the back yard.
The sterba is hung directly over the top of my house.
the bottom element of the array is up at 35'.
The design frequency for maximum gain was 28.5mhz
THIS DESIGN IS FOR ALL BAND USE.... 160 THRU 10 using a Tuner!.
The standard formulas below were used
468/freqmhz = 1/2 wavelength section
234/freqmhz = 1/4 wavelength section.

My antenna elmer is W1VDE, Roger, He was part of the team that built the dishes at Arecibo, Puerto Rico years ago. I am attaching the other half of my antenna system(John is refering to the Tunable Radial System in pic below) for your review and comments. This system adds approx. 3db to 6db of additional gain to the Sterba.
If you need further information please do not hesitate to call.
Hams that have installed the curtain have liked it so much they have put up more than one."
73

John / N0KHQ / St. Louis
Always on 17M
Antennas:
You can build 'um better than you can buy 'um.

This taken from recent email from John, N0KHQ
Hi Don,
Okay, here we go. Here's an example for 17 meters.
The Sterba is approx. 50' long x 16' high.

Example Gain on 17m:
17m squared / 12.56 (4pi) = 23 sq. meters of aperture

50' / 3.28 (1 meter) = 15.24 meters
16' / 3.28 (1 meter) = 4.878 meters

15.24m x 4. 878m = 74.34 sq.meters

74.34 / 23.00 = 3.24 log x 10 = 5.09 db + 1.64 (gain over dipole) = 6.73 db gain on 17m.

If you are running 150w x 6.73 = 1009.5 watts of effective radiated power output on 17m.

On 10m (the design frequency) the gain is 10db +/-
The above formula can be used for calculating aprox gain on other bands

I need to clear up something that I may not have explained very well. This design allows a ham to use this array on all bands 10m through 160m, on 160m the swr bandwidth is pretty narrow.. There is no need to build arrays for different bands.
I have not tested other Sterba's designed for other bands as to their performance from 10m through 160m.

ABOUT FEEDLINE LENGTH
The reason the 450 ohm feed line is 130' is because it is about 1/4 wave length on 160m which is a high current point. So, what ever is tied to the end of the feed line.....is going to get out!.
For the feed line length (130') I have tried different lengths but 130' seems to be the minimum optimum length to get on 160 m.
Example:
I used 935/1.931=484.2' / 4 = 121.0' . This is a 1/4 wavelength at 1.931mhz, in the example. Some have said that 450 ohm ladder wire is actually 400 ohm. Some have said the velocity factor is not .95. Man, who really knows?

This antenna serves me well. Since I only have room for one.
A good thing to remember is the phrase "Volumetric Efficiency".

I have sent this design to a number of hams, the design has made it all the way to Europe. If you would like to contact a friend of mine that is running this array his call sign is KB4CCM, Bill. He liked it so well he put up two!

(Bill, KB4CCM WRITES THIS ABOUT THE DESIGN..... TAKEN FROM AN EMAIL)

"If hams interested in the Sterba wish to contact me, (see email address below)
I would be glad to answer any questions that I can and
give tips on construction. As John has told you, I

have taken down all of my other wire antennas (dipole, extended double Zepp and Half square) due to the performance and versatility of the all band Sterbas.

By the way, you may be interested in an article I wrote which is published in the ARRL Antenna Compendium # 7 now on sale. It is entitled 'The ODD QUAD' and details instructions on how to build a strong and cheap (using locally available materials) two element Quad that can be installed and rotated without a man made vertical structure. This Quad, along with the Sterbas, may be the only antennas I will ever need or use for the rest of my ham career."".....Bill KB4CCM

BACK TO THE PROJECT

The buried radial system that is used with the Sterba adds approx. 3db to 6db of gain. Once you add this gain to the gain of the curtain, the total gain of the antenna system is 9db to 12db of gain on 17m.

A good reference book that is available that will help hams to understand the importance of a good buried radial system is "Low Band DXing" by ON4UN cost $30.
Hams running the low bands are experts on antenna systems.
I call it my antenna bible.
Hope all this helps.....my head hurts!!!!!!!
73
John / N0KHQ / St. Louis
Always on 17M
Antennas:
You can build 'um better than you can buy 'um.

Editor's note:
Many thanks to John, N0KHQ for sharing this information for all Hams to enjoy!
and to Bill, KB4CCM for standing by for any questions you may have.
If you have questions about this project please contact them via email at:
John, N0KHQ Email me here
Bill, KB4CCM Email me here

NOTE: THIS DESIGN is for ALL BANDS ---160 thru 10 Meters using a Tuner!

BUILD A STURBA CURTAIN ANTENNA FOR ALL BANDS!

The Broadside Curtain Aray

THE TECHNICAL STUFF

GET OUT THAT CALCULATOR!

Back to Antenna Lab

(See "About feedline length" at bottom of article)

Antennas & More for HF antennas

NEW!
See John's
17 Meter Coaxial Moxon.
40 % reduction in size over standard Moxon!
Click here

RADIAL SYSTEM INFO

The Sterba Curtain, as designed is non resonant.
There are a couple of bands where the swr will drop to 3:1.

The installation of the tunable buried radial system will help on 160m, 80/75m and 40m.
If I were you I would put out radials for those bands only.

4 for 160m
4 for 80/75m
4 for 40m

Remember, the radials should be cut at .15 of a wave not .25. There is a 10% shortening factor when radials are either on the ground or buried just below the grass level. By installing the radial system you will add 3db to 6db of gain to the gain on the curtain now. (ref. ON4UN - Low Band DXing)

I bought a HedgeHog from Lowes to dig the trenches $50 and used, what I call chain link fence wire ties, the aluminum wire ties are already bent in the shape of a "U" and hold the wire down very nicely.

You wont need to run an amplifier with the antenna for good performance.

"Curtains for You"

As Forty Meters has traditionally been the most productive band on CW for our Club Field Day efforts, I would like to put up more than the usual dipole this year. The most attractive antenna appears to be a Sterba Curtain. I would ideally like the antenna design to center on 40, but this may not be practical. Perhaps we could design it for 30 Meters- that would be close enough to tune it and enjoy most of the benefits and gain of the Sterba.

My plan would be to put up the 2 section Sterba, run ladder line to a tuner in a temporary "dog house" to protect the antenna tuner in it, adjust the tuner at the center of the 40 Meter CW band, and run the needed length of feed line with coaxial cable to the Transceiver. Once tuned, we should be able to operate the whole 40 meter CW band without having to retune.

A two-section Sterba would be a full wavelength from end to end, and a half wavelength in height. Ideally, the bottom should be a half wavelength off the ground. At lower heights, it still has to be a lot better than a dipole. The 2 section Sterba should fit nicely between the stadium light poles. Provisions could be included to rotate the array.

Required materials would include:

1. A full wavelength of 450 ohm ladder line. For 40 meters this would be approximately 133 feet. For 30, this would be about 90 feet. Added to this would be a short section to the dog house.
2. 3 full wavelengths of wire. For 40 meters, about 400 feet, for 30, about 300 feet. I have been collecting wire, this should be no problem.
3. 8 Insulators (can be fabricated)
4. Tuner and Coax.
5. Rope- undetermined amount.

The Sterba Curtain: What is it, and what makes it so great?

The Sterba is a gain antenna, with the gain depending on the number of dipoles pairs. The unique signature of the Sterba is the use of ladder line with a half turn to provide the proper phase to the dipoles.

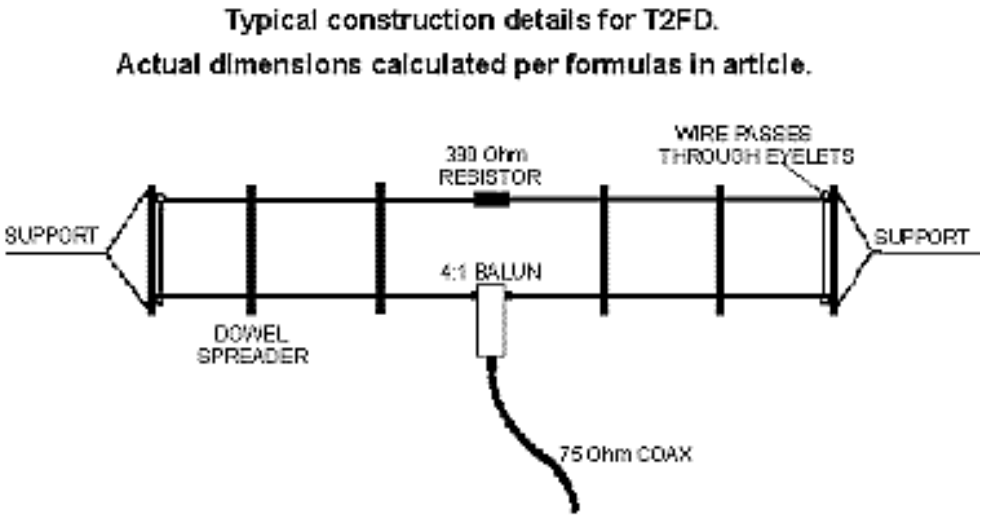
The basic Sterba unit is 2 parallel half wave dipoles with a half wave phasing harness between them. You may add as many units as you have wire and room.

The amount of wire in this antenna alone makes it a great receive antenna.

March 2002 KB8PGW

hcdx wire antennas

More on the T2FD
- Main
- Design
- Feeding
- Compare
- Opinion



The formulas for calculating T2FD dimensions are as follows.

1. The length of each leg ("A") from the center is equal to 50,000 divided by the lowest desired operating frequency (in kHz) and then multiplied by 3.28. The answer is in feet.

2. The spacing between radiating wires ("B") is equal to 3000 divided by the lowest desired operating frequency (in kHz) and then multiplied by 3.28. The answer is in feet.

3. The sloping angle for a non-directional pattern should be on the order of 30, but 20–40 is acceptable.

Example:
To design a T2FD for the center of the 90 meter band (3300 kHz) and up:

	Desired frequency	Spacing
Leg "A"	(50.000 / 3300) x 3.28	49.70 feet
Leg "B"	(3000 / 3300) x 3.28	2.98 feet

Total length of the antenna would be 99.4 feet (2 x 49.7), and the width would be 2.98 feet ("B").

The total wire used to complete the loop equals 204.76 feet (4 x 49.7)

- Front page
- DX News
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Open links in new Window

OPERATING MODES

QUICK FIND

AM	GPS	Meteor Scatter	Pedestrian Mobile
AMTOR	GTOR	Microwave	PSK31
APRS	HandiCapped Ham	MMTTY	QRP
ATV SSTV	Hell Modes	Mobile, Auto, RV	RTTY
Bicycle Mobile	HF Digital	Modes Of Operation	Satellites
CLOVER	HF DX	Modulation	Software
CW	HF Packet BBS	Morse Code	Spread Spectrum
Digital	HF Radio	Motorcycle Mobile	SSB
ECHOLINK	HF Portable	MFSK16	SSTV
Emergency Radio	ILINK	MT63	THROB
EME, SETI, Astronomy	IRLP	NAVTEX	T-Hunts
FAX	Laser	PACKET	VHF DX
FM & PM	Low Freq	PACTOR	WEFAX
			WIRES II

MODES OF OPERATION THESAURUS

How Modulation Modes Work -- Very Visual Explanations for AM, FM, SSB, DSB, QAM

Telecommunications Glossary

Modulation Codes -- A1A, etc

AMPLITUDE MODULATION

The AM Window Pages -- Audio, Events, Freqs, Nets, Resources, Technical

The AM Window --Loaded with goodies for the AMer

AM Expained

CW AND MORSE CODE

Go To The CW Page -- 20 Links to CW Trainers, Aids, 200 Practice QSO's, History and Much More

---

Handi-Hams -- Handicapped, Visually Impaired

Courage Handi-Ham System

Power/SWR Meter For The Sightless -- From RF Applications

ARRL Amateur Radio for the Blind -- Many Articles

ARRL SourceBook for the Disabled

GPS System For The Sightless

Wheelchair Accessible Minivan, the Vision™.

HF MOBILE

Mobile Mail List Reflector -- Subscription Info

HF Mobile Of The Month -- Via K2BJ

HF Mobile Message and Discussion Board -- From K2BJ.COM

HF Mobile -- A San Diego Mobile -- From KA6WKE -- Great Installation Tips

Great Mobile Installation Tips -- From Alan Applegate, KØBG Via eHam.Net

Home of HF Mobile on the World Wide Web! By K2BJ

Automotive Noise Elimination By Stuart Downs. WY6EE -- A Must Read

Solving Ignition Noise RFI - Via The ARRL

Noise Suppression Techniques - From K2BJ

Grounding Concepts and Techniques - From K2BJ

Automobile Articles - Via The ARRL

NX7U Mobile Pages -- Check This Out

From The News Groups:
Someone asked me to post the following summary of three California 75m Mobile Shootouts:

Bugcatcher/Screwdriver with top hat	**0 dB reference**
Bugcatcher/Screwdriver, no top hat	**-3 dB, -50%**
Hustler	**-7 dB, -80%**
Outbacker	**-9 dB, -88%**
Hamstick	**-12 dB, -94%**
Whip with autotuner	**-14 dB, -96%**

Considering that the top legal 75m mobile antennas are only 5%-10% efficient to begin with, a Hamstick radiates ~0.6 watts in the far field with 100 watts input. 75m Hamstick users must really enjoy QRP. :-)
--

73, Cecil http://www.qsl.net/w5dxp

MOTORCYCLE, BICYCLE, AERONAUTICAL, AND PEDESTRIAN MOBILES

Motorcycle Amateur Radio Club -- MARC

Motocycle Mobile Forum -- From eHam.net

Motorcycle Special Events Team -- Texas Style

Bicycle Ham Pages

Bicycle and Pedestrian Mobile Pages - Via HFPack - The Portable Group

Pedestrian Mobile - From W3FF

Aeronautical Mobile Page

HF PORTABLE

Goto The HF Portable Pages

HF RADIO

HFRADIO.ORG -- Propagation, DX, Callsign/QSL routes, Swap and Shop, Online Map Creation, Learning / Licensing -- By NW7US,Tomas

---

LOW FREQUENCY OPERATION

LOWFERS Longwave Club of America. -- The World of Radio Below 500 kHz

Low Bands -- 160, 80, 40, and 30 meters From K3KY

160 Meter Mail List Reflector -- Subscription Info

QRP OPERATION

Go To The QRP Page

HF DIGITAL AND PACKET RADIO

What Digital Signals Sound Like -- Many many digital signal sounds -- From World Wide Utility News

Another Digital Modes Sample Sounds -- listen to the sounds -- from KB9UKD

FYI: GENERAL DIGITAL MODES DESCRIPTIONS - On Line Sounds

ALSO SEE PSK31 SOFTWARE

Digital Signals FAQ -- Loaded -- From WUN

About Digital Ham Radio -- From NB6Z

AMTOR From NB6Z

CLOVER -- From Hal Communications

CLOVER -- Jim's Digital Gazette By N2HOS

GTOR From NB6Z

G4GUO's Charles Brian digital SSB protocol. From G4GUO

G4GUO's Charles Brian digital SSB protocol. -- AOR Version From KQ6EH

HELL Modes From NB6Z

IRLP, ILINK, ECHOLINK, WIRES II

MFSK16 From NB6Z

MT63 From NB6Z

THROB From NB6Z

Packet Software -- Several to choose from

Tucson Amateur Packet Radio -- TAPR, All About Packet, APRS, DSP, Kits & projects

North American Digital Systems Directory -- From TAPR

Superb Packet Tutorial -- by Larry Kenney, WB9LOZ

Packet Radio Training Course -- From The Ventura County ARES AREA 6 Training Group

Another Great Packet Tutorial -- by Lincoln ARC

Still Another FB Packet Tutorial -- Buck Rogers K4ABT

Great Packet Listings -- From The QRZ Folks

Packet dot Com -- Loads of Packet info

US Packet Net -- Lots Of Packet Info

PacketZone -- From Jan, DG1LJP - portal for Packet Radio-Users and offers a lot of links.

Packet Glossary -- The Original By Buck Rogers K4ABT

TAPR & Packet Mail List Reflector -- Subscription Info

Broadcast protocol in packet radio. -- Via TAPR

PACTOR NEWS -- Phil Sussman - N8PS

PACTOR Without A TNC -- From Tom Sailer, HB9JNX

PACTOR-II -- SCS Home Page -- The Works - includes a FB FAQ

Jim's Digital Gazette By N2HOS -- All about digital modes--RTTY, Amtor, Pactor, Clover, G-tor

HF Digital Radio -- Well Done Tutorials and Operating Info -- By NB6Z

HF Packet BBS's -- Try The Following Frequencies
14,105.5, 14,095.2, 10,149.25, 10,147.25, 7091.5 and 7103.5

Tucson Amateur Packet Radio Corp. -- THE definitive Source

North American Digital System Directory -- BBS, DX Packet Clusters, Nodes, Maps

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AX.25 Amateur Packet-Radio Link-Layer Protocol --Via TAPR

Great Packet Page -- By G. E. "Buck" Rogers Sr K4ABT

Portable Packet Web Page ! -- By N1RWY -- Includes Baycom version 1.4 and Baycom version 1.5

MMTTY - Sound Card RTTY!!

MMTTY - program

RTTY Reflector -- Subscription Info

RTTY Software

RTTY Fan Club - Lots Of RTTY Info From Sotiris SV1BDO

ARRL RTTY Transmissions -- Schedule

RTTY DX News -- From Jim's Gazzette

RTTY HINTS -- From Kantronics

RTTY Loop -- By Marc I. Leavey, M.D. WA3AJR

RTTY Pages -- Including DXing, Contesting, Software Reviews -- From N1RCT

PSK31

SEE SOFTWARE FOR PSK31

Has Awards and General PSK31 Pages as Well

APRS San Diego -- Via SANDRA

APRS Primer - Via SANDRA

APRS Station Locations - findU.com -- A must see -- shows APRS Stations, maps, position, much more, all across the USA!

APRSPoint -- Mapping Software Program -- From KF6ZDM --- With APRSPoint, you don't have to scan maps any more!!!

APRS Mail List Reflector -- Subscription Info

- **APRS -- Automatic Packet Position Reporting System -- Lots of Info & Links**
- **APRS -- From TAPR**
- **APRS -- Tri-State APRS Working Group**
- **APRS -- Kansas City APRS Working Group**

GLOBAL POSITIONING SYSTEM

Yogi Berra Once Sed To His Wife
"I think we're lost but we're making good time"!

GPS Made Easy : Using Global Positioning Systems in the Outdoors by Lawrence Letham
Buy the Book Today!

GPS Land Navigation; A Complete Guidebook for Backcountry Users of the NAVSTAR Satellite System by Michael Ferguson, Randy Kalisek, Leah Tucker
Buy the Book Today!

A Comprehensive Guide to Land Navigation with GPS by Noel J. Hotchkiss
Buy the Book Today!

GPS NOW 10 TIMES AS ACCURATE -- Need URL Explaining this -- anyone??

The Global Positioning System -- By Robert A. Nelson

GPS -- From USNO GPS Timing Operations

GPS Primer -- From The AeroSpace Corp.

GPS Mail List Reflector -- Subscription Info

GPS -- Global Positioning System Overview-- From Peter H. Dana Department of Geography, University of Texas at Austin

GPS Time Series -- From NASA

NAVSTAR GPS Operations -- From The Time Service Dept., U.S. Naval Observatory

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SLOW SCAN TV

SSTV & ATV Mail List Reflector -- Subscription Info

MMSSTV - Sound Card SSTV!!

SSTV -- From Wayne's World -- Click on "Ham Radio" -- Lots of Programs & info

SSTV -- Excellent Slow Scan TV Page -- From PA3GPY

G0ITP's SSTV Page -- All About SSTV -- Numerous Links

W5NCD's SSTV Web Page -- Well Done SSTV Pages

FMSSTV - mostly about SSTV, slanted toward FM 2 meter -- From Guy Clark, AB0DP

Radiofax, Wefax, Pressfax, Photofax and Amateur, SSTV -- By Marius Rensen

ON4JU SSTV Website

ALSO See SSTV Software -- Several To Choose From

FAX & WEFAX

- **World Wide Marine Radio Facsimile Broadcast Schedules -- From NOAA**
- **National Weather Service (NOAA) -- Marine Home Page**
- **WEFAX Frequencies**
- **WEFAX -- From Kantronics**
- **WEFAX - Mscan Meteo -- Program for high resolution weatherfax reception.**
- **Radiofax, Wefax, Pressfax, Photofax and Amateur, SSTV -- By Marius Rensen**

METEOR SCATTER

- **High Speed CW Meteor Scatter Pages**
- **Operating Meteor Scatter Mode -- From The Six Meter World Wide DX Club**
- **WSJT Meteor Scatter Programs -- by K1JT**
- **Working DX on a Dead 50MHz Band Using Meteor Scatter -- Palle Preben-Hansen, OZ1RH**

FAST SCAN TV -- ATV

- **ATV From France -- ATV repeaters in south of France, frequencies, modulations, used in France and in Europe -- From André F5AD**
- **Amateur Television and High Speed Data Signaling From F4DAY**

Amateur Television Network -- Atlanta

ARIZONA AMATEURS ON TELEVISION From KB7TBT

British Amateur Television Club -- With Amateur TV related software available for download

IK1HGI PAGES - TELEVISION AMATEUR ATV

Klamath Amateur Television -- Southern Oregon ATV

OH3TR ATV Pages -- Many Excellent Links and Tech Data

VE7RTV ATV Repeater -- British Columbia Television Group Website

San Francisco Bay Area ATV -- WA6ZJG

LASER COMMUNICATIONS!!!

Lasers -- K3PGP Experimenter's Corner From VLF to Light

Amateur Radio Laser Communications -- From Jim Moss, N9JIM/6

Laser Mail List Reflector -- Subscription Info

MICROWAVE & SPREAD SPECTRUM

SAN BERNARDINO MICROWAVE SOCIETY -- DEDICATED TO THE ADVANCEMENT OF COMMUNICATIONS ABOVE 1000 MHz

Microwave Page From WA1MBA -- Devoted to Amateur Radio Microwave interests including the MICROWAVE REFLECTOR.

- **Microwave -- RSGB Microwave Committee - Loads of uWave Goodies**
- **Microwave -- The World Above 1000MHz From G3PHO Peter Day Sheffield**

- **Spread-Spectrum Radio Design -- Via TAPR**
- **Spread Spectrum -- TAPRs Spread Spectrum Communications Page**
- **Spread Spectrum Scene - Online -- By KC6YJY, Randy Roberts**

VHF DXING

VHF Mail List Reflector -- Subscription Info

North East Weak Signal VHF Group -- The Works On VHF, Contests, Grid Square Maps, Calculate Your Grid Square

Hawaii Tropo Ducting -- From KH6HME

TRANSMITTER HUNTING

- **Jerry Boyd's T-Hunt Web Site**
- **Hudson Valley DFA -- Includes Home Brew Projects**
- **Homing In -- From KØOV**
- **DOPPLER D/F INSTRUMENTS -- From WB6EYV**
- **Stalking the elusive "T" -- Southern California Transmitter Hunts**

Doppler Systems Inc. -- Designs and manufactures radio direction finding equipment for amateur use

MicroFinder Doppler RDF -- From AHHA! Solutions

NAVTEX

Note that some Ham TNC's and Software are capable of receiving NAVTEX Broadcasts. See manufacturers pages and software pages. These broadcasts are easily received and decoded provided you are within range of a NAVTEX station and have a suitable terminal or decoding software, listen on 518KHz. The transmissions are short, perhaps 5 to 10 minutes, repeated 4 hourly.

NAVTEX -- Loads of info by Cor van Soelen

AC6V's Amateur Radio Links **GO BACK TO THE BIG INDEX**

AC6V's Amateur Radio Links **GO BACK TO THE QUICK INDEX**

AC6V BRAG TAPE

Last Update: September 02, 2004

TELREX ANTENNA TRAP REPAIR -- FROM N6KI

Rod,

I just wrote up this Telrex antenna Trap repair procedure for Bob, K6XX. If you have room somewhere on your web page maybe you could include it somewhere so other Telrex owners around the world can keep em ticking!

73, Dennis N6K

Hi Bob, Yes Telrex traps do tend to toast but not since I modified the 15 and 20 mtr traps on the driven element of my 7 element Monarch which covers 40 thru 10 mtrs. Never had a problem with the 10 mtr traps. I marked where the driven elements assemblies were inserted into the short stub coming off the boom and was able with my tower cranked down and another person on my patio roof supporting and pushing up on the ends of the driven element, to remove the 2 sides. (I borrowed my neighbors pool cleaning poles and mounted a donut shaped piece of aluminum to the end of the pole and had my friend capture the end of the element so he could lift and take the weight off my end near the boom).

The failure mechanism is that even though the doorknob cap used in the trap is rated at 5 to 7 kV ACROSS the cap, Telrex, in their infinite wisdom, had laid the doorknob cap directly on top of and touching the coil section of the trap! So, the caps appear to flash right up through the SIDE of the cap from the coil! The object of this repair exercise is to remove the old burned out doorknob cap but replace it with 2 doorknobs in series, allowing you to lift the straps that the ends of the doorknob cap is attached to. Two advantages of this type of repair are:

1. You now have double the voltage protection across the two 5 kV doorknobs caps.

2. The doorknob caps no longer lie directly on top of the trap coil. Now, on my Monarch, the value of the doorknob happened to be 33pF (Since they seem to burn one side of the driven element at a time, I was able to chisel off the goop on the corresponding good trap on the other side of the driven element and read the value, maybe yours are different value.) Ok so here's the drill on extracting the old burned doorknob and then resealing the trap. Once you get the element on the ground I went to a machinist (who happened to be my father) and he made me a small sharp chisel shaped piece of metal. I used it and a small ball peen hammer and I CAREFULLY chiseled off that brownish colored goop used to seal the the trap. I exposed the doorknob capacitor (or what was left of it) and I also exposed the flat metal straps on each side all the way down to and including the nut and screw end of the straps.

Be very carful when chiseling around the coil not to dig the end of the chisel INTO the coil or flat

metal straps. If you do hit the coil by accident, although you won't cut through the coil you will nick it pretty good. Take a file to it later and file off any sharp edges or burrs you create. Using a small wire brush, be sure to remove ALL traces of the black burnt carbon and other debris that the burned out doorknob capacitor deposited in and around the coil.

You may need a small chisel tool to clean in between the coil windings. Since it's hard to find the exact 32.XX to 33.XX pF cap I replaced the original cap with 2 doorknob caps in series using a short piece of threaded rod to screw then together. (Changing the value of the combo of caps by just a few tenths of a pF moves the resonant point quite a bit. 0.5pF too much or too little could move you 150 kHZ if I remember right!) I found plenty 50 to 75 pf 5 kV doorknob caps on the surplus market and put 2 caps in series of various values and was able to fine tune my antenna resonance points to exactly which part of the band I wanted. (This is NOT the fun part, as I started with some series combination value around 32.5 pF using my hand held capacitance meter, then, before sealing the traps, I reinstalled the driven element and cranked the tower up and measured resonant frequency.)

Changing the value of the combo of caps by just a few tenths of a pF moves the resonant point where I wanted it. I originally tried to use a grid dip meter t0 measure the resonant freq of the good corresponding trap on the opposite side of the driven element but that didn't work out for some reason (It's been a few years since I did the repair.) For some reason and maybe because I lifted the caps off of and away from the coil, my notes indicate that I wound up using combinations of caps that yielded capacitance in the 30 to 31.5 pF range, nowhere near the original 33 pF (marked value) they replaced!!

The opposite ends of the dual doorknob caps assembly are now connected to each strap and now it is easy to position the caps at least 1/4 inch away from the coil. Once you get the antenna resonant where you want it, seal the repair by first wrapping the coil from end to end with Kapton Tape which should be available from industrial electrical supply companies. I utilized a 1/2' width tape and applied 2 layers to the coil using an overlapping wrap. Then apply Dow Corning 738 Electrical Sealant to cover all exposed areas of the coil and doorknob capacitors. 738 is UV stable and has a dielectric property of 400 to 500 Volt per 0.001 inch. That will give you plenty of arc protection now that you have the caps 1/4" off the coil! (The sealant did NOT affect the resonant point, luckily!)

Unfortunately 738 only comes in white which makes for a very ugly repair, but up at 50 ft or more, who's going to see it except the birds! It also has only a 6 month shelf life so a lot of places won't stock it and you may have to special order it. There's an industrial supply company in San Diego called Yale Enterprises (619 299-7710) that used to sell it in 3 OZ small tubes but you may have to buy the caulking gun size. I believe that one 3 OZ tube will easily have enough sealant to redo 2 traps. The small tubes used to cost @ $7 and @$20 for the caulking gun size. There's also a place in San Jose area called Anderson Supply that carries it but may make you buy the larger amount to meet minimum order. Caveat Emptor if you use any other type of sealant, make sure it's UV impervious, temperature stable and has enough electrical dielectric strength, voltage wise.

Also, a different type of sealant might shift the resonant point !!! After you repair the burned trap,

go back and chisel out the remaining good trap doorknob and replace it or I can guarantee that in the dead of winter when your tower is covered with ice, the other side WILL fail. My Telrex Monarch continues to provide good reliable service, contest after contest on 40 thru 10 meters, pumping 1.5 kW into it for 48 hrs at a time! (I just found a note showing I repaired it 8 years ago!) If this 200 lB beam wasn't such a killer antenna on 40 mtrs with 3 almost full size elements, I probably would have not gone through this much trouble to repair it. BTW, it is was originally rated for 100 MPH wind and 5 KW and cost over $2000 in 1982. So who says a trapped antenna can't compete!

73, Dennis Vernacchia N6KI / ZF2AR

AC6V's Amateur Radio Links **GO BACK TO THE BIG INDEX**

AC6V's Amateur Radio Links **GO BACK TO THE QUICK INDEX**

Last Update:July 31, 2002

AA3RL Transmission Line Calculator

Introduction:

There are a variety of programs available to the Radio Amateur that will calculate virtually every transmission line parameter that one may need. However, I have found the routines to be inconveniently scattered among many different programs. In addition, the programs usually require the data to be entered in serial form in order to obtain the calculated output. If another calculation is desired, all the data must be re-entered. As a result, I felt the need to create a spreadsheet style program where all input values are displayed and multiple "what-if" data changes may be implemented quickly with the click of the mouse.

Initially I used the spreadsheet to design Stub and Series Section Line Matches. Since then other utility functions were added that has enabled it to replace the Smith Chart, the pocket calculator, and the many computer programs for transmission line design applications. I hope you will find it equally useful.

Caution:

All formulas assume lossless transmission lines, but these results are usually within the measurement tolerance achievable by the amateur in normal situations. However, as line loss increases, i.e. with high SWR or long transmission line lengths, these approximations will diverge from the actual values. An excellent program to check the results for lossy lines is TLA, which is included with recent versions of the ARRL Antenna Book.

Format:

The spreadsheet is divided into convenient functional areas. Each area performs one of the calculations listed below. All sections share a common layout. The input data is grouped on the left side of the page, and the output data is grouped to its immediate right. Output values are shown in bold type. On the far right side of each section there are areas that are used for computational purposes, in order to break complex formulas into more manageable chunks. If these cells are disturbed or over-written the calculations may be cease to function properly. I have learned from experience to keep a backup copy of the spreadsheet for when this eventually happens.

Contents:

The program performs the following calculations:
Length Conversion: Wavelength, Degrees, Feet
Series and Parallel Equivalent values
Capacitance and Inductance required for given reactance (X)
Air Wound Coil construction
Reactance and Length of Stubs
Stub Matching - Impedance (Z) over a 180 degree cycle along a lossless transmission line.
Impedance (Z) at any single point on a lossless transmission line (for stub matching)
Series Section Line Transformer (analytic solution) for impedance matching

Length Conversion: Wavelength, Degrees, Feet

In the screen shot below, we want to make a transmission line 1/8 (0.125) wavelength long, using 50 ohm line with a velocity factor of 0.78, and operating at a frequency of 7.0 MHz. This data is entered on the left side. The output indicates that the length should be 13.7 feet, which is 45 degrees. Alternatively, the reverse calculations may be performed.

TLCalc1.xls

	A	B	C	D	E	F	G
3	**AA3RL - TRANSMISSION LINE CALCULATOR**						
4							
5	**Length Conversion: Wavelength, Degrees, Feet**						
6							
7	**INPUT DATA:**			**OUTPUT DATA**			
8	Wavelengths:	0.125	------>	Length (Feet):	**13.70**		
9	Freq (MHz.):	7.000		Degrees:	**45.00**		
10	Vel Factor:	0.78					
11							
12	Degrees:	56.000	------>	Length (Feet):	**20.76**		
13	Freq (MHz.):	7.000		Wavelength:	**0.156**		
14	Vel Factor:	0.95					
15							
16	Length (Feet):	13.700	------>	Wavelength:	**0.12**		
17	Freq (MHz.):	7.000		Degrees:	**45.0**		

Sheet1 Sheet2 Sheet3 Sheet4 Sheet5

Series and Parallel Equivalent values

Series and Parallel equivalent circuits are very useful for transmission line matching, as well as other circuit design applications. Conversions may be made in either direction.

rial 10 B I U $ % , B46 34

TLCalc1.xls

	A	B	C	D	E	F	G
19							
20	**Series and Parallel Equivalents:**						
21							
22	**INPUT DATA:**			**OUTPUT DATA**			
23	Series Impedance Z (R +/- jX)			Parallel Equivalent (R +/- jX)			
24	R(s)	77.50	------>	R(p)	**134.39**		
25	X(s)	-66.40	------>	X(p)	**-156.86**		
26							
27	Parallel Impedance Z (R +/- jX)			Series Equivalent (R +/- jX)			
28	R(p)	425.00	------>	R(s)	**25.00**		
29	X(p)	-106.25	------>	X(s)	**-100.00**		

Sheet1 Sheet2 Sheet3 Sheet4 Sheet5

Capacitance and Inductance required for given reactance (X)

At any given frequency, input the desired inductive reactance (+X) or capacitive reactance (-X). The spreadsheet calculates the required inductance (uH.) or capacitance (uF and pf).

32						
33	**Capacitance and Inductance required for Reactance = X**					
34						
35	**INPUT DATA:**			**OUTPUT DATA**		
36	Freq (MHz.):	3.50				
37						
38	Reactance, Ind (+X)	800	------>	Induct (uH)	**36.38**	
39						
40	Reactance, Cap (-X	500	------>	Capac (uF)	**0.000091**	
41				Capac (pF)	**91**	
42						

Reactance and Length of Stubs

Another way to obtain a desired reactance is with the use of transmission line stubs. Input the desired inductive reactance (+X) or capacitive reactance (-X). The spreadsheet calculates the required length of stub: open for -X, shorted for +X.

Alternatively, given a stub of any length in feet which is less than ¼ wavelength, the reactance will be calculated. Results may not be accurate if the electrical length is greater than 90 degrees.

TLCalc1.xls

	A	B	C	D	E	F
31						
32	**Reactance of Transmission Line Stubs**					
33	*** **Length MUST** be less than 1/4 wavelength (90 degrees) to be accurate					
34						
35	**INPUT DATA:**			**OUTPUT DATA**		
36						
37	Stub Data:			Reactance:		
38	Stub Length (Feet):	13.700	------>	X (open)	**-50.00**	
39			------>	X (shorted)	**50.00**	
40	Freq (MHz.):	7.000		Length (Deg)	**45.00**	*Result NOT accurate if Length > 90 deg*
41	Vel Factor:	0.78				
42	Z (line), resistance:	50				
43						
44	Desired Reactance:			Stub Needed:	Degrees	Feet
45	Capacitive Reac (-X)	34	------>	Length (open)	**55.8**	**5.66**
46	Inductive Reac (+X)	34	------>	Length (shorted)	**34.2**	**3.47**
47	Freq (MHz.):	21.000				
48	Vel Factor:	0.78				
49	Z (line), resistance:	50				
50						
51						

Sheet1 | Sheet2 | Sheet3 | Sheet4 | Sheet5 | Sheet6 | Sheet7

Stub Matching

Using tuning stubs to match an antenna and transmission line is quite popular and useful. While the theory of stub matching is beyond the scope of this article, the following example illustrates how it may be performed using the AA3RL TLCalc spreadsheet.

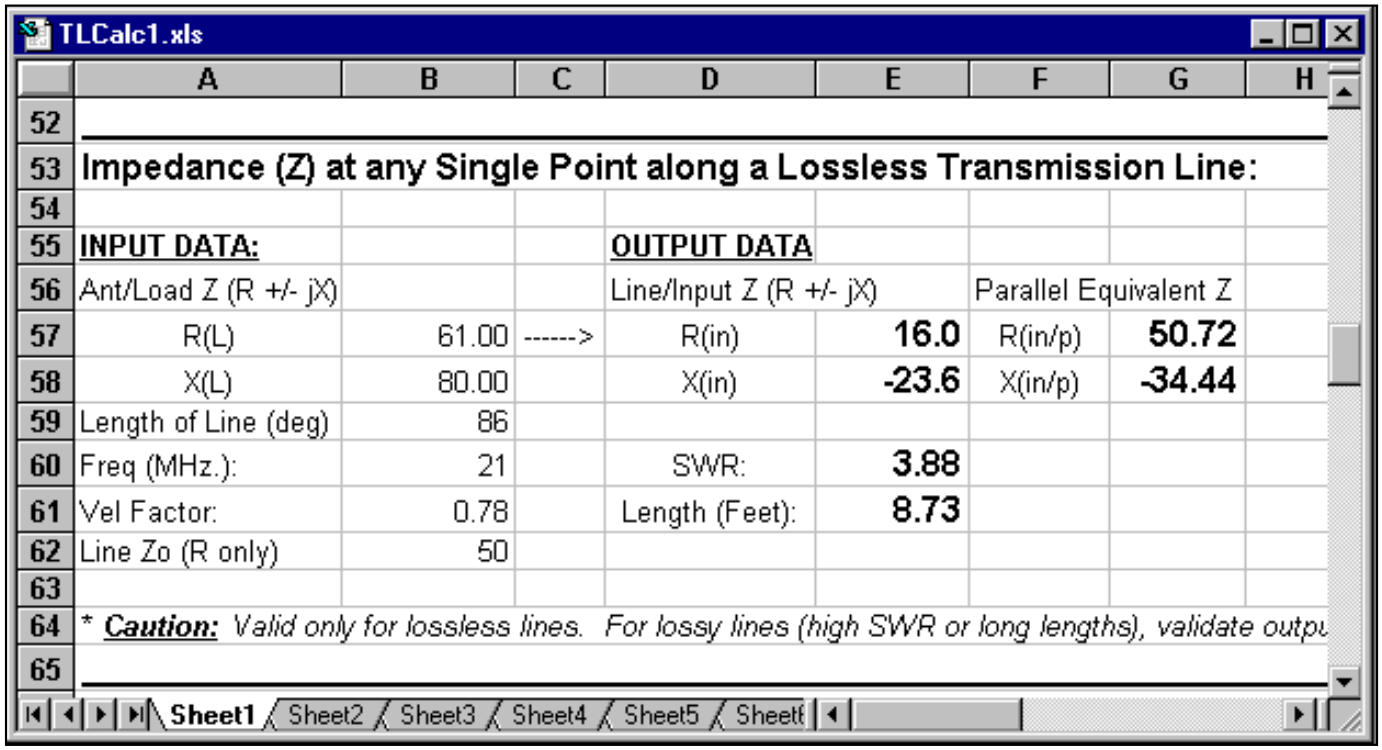
We wish to match a 50 ohm feedline to an antenna with Z = 61 + j80. These values are entered as inputs. We immediately notice that the SWR is 3.88. Next, scan the third Column of the OUTPUT DATA to find where the parallel equivalent resistance, R(p), approaches 50 ohms. We see that this occurs somewhere between 80 and 90 degrees. A good initial guess would be 85 degrees. We also note that R(p) is not changing too quickly at this point so that measuring the stub insertion point will not be overly critical.

TLCalc1.xls

	A	B	C	D	E	F	G	H	I	J
66	**Impedance (Z), over a 180 degree cycle, along a Lossless Transmission Line:**									
67										
68	**INPUT DATA:**			**OUTPUT DATA**						**Comp I**
69	Ant/Load Z (R +/- jX)									A(swr)
70	R(L)	61		SWR:	**3.88**					B(swr)
71	X(L)	80								
72	Line Zo (R only)	50								
73										
74	**OUTPUT DATA**					Parallel Equivalents		**Computational Informa**		
75	Length of Line (deg)			Line Z (R)	Line Z (X)	R(p)	X(p)	Len (rad)	R num	R/X den
76	0			**61.0**	**80.0**	165.9	126.5	0	61	1
77	10			**112.0**	**90.2**	184.6	229.3	0.174533	62.897	0.56162
78	20			**185.9**	**37.5**	193.5	959.7	0.349066	69.081	0.3716
79	30			**162.0**	**-69.1**	191.5	-449.2	0.523599	81.333	0.50195
80	40			**89.2**	**-89.4**	178.9	-178.4	0.698132	103.95	1.16531
81	50			**50.3**	**-73.3**	157.2	-107.8	0.872665	147.64	2.93624
82	60			**32.1**	**-55.8**	129.0	-74.2	1.047198	244	7.60264
83	70			**22.9**	**-41.4**	97.7	-54.1	1.22173	521.47	22.768
84	80			**17.9**	**-29.7**	67.2	-40.5	1.396263	2023	113.062
85	90			**15.1**	**-19.8**	41.0	-31.3	1.570796	2E+34	1.1E+33
86	100			**13.5**	**-10.9**	22.3	-27.7	1.745329	2023	149.359
87	120			**13.1**	**5.6**	15.4	36.2	2.094395	244	18.6878
88	130			**14.0**	**14.0**	28.0	28.0	2.268928	147.64	10.5635
89	140			**15.9**	**23.2**	49.7	34.1	2.443461	103.95	6.53555
90	150			**19.4**	**33.7**	77.9	44.8	2.617994	81.333	4.19699
91	160			**25.6**	**46.2**	109.2	60.4	2.792527	69.081	2.70101
92	170			**37.2**	**61.8**	139.7	84.2	2.96706	62.897	1.69012
93	180			**61.0**	**80.0**	165.9	126.5	3.141593	61	1
94										
95	* ***Caution:*** *Valid only for lossless lines. For lossy lines (high SWR or long lengths), validate output data with TLA*									
96										

Next move to the spreadsheet section for computing the impedance at a **single point** along a transmission line. Enter the same

inputs: Z = 61 + j80, Frequency = 21 MHz., TL = 50 ohms, velocity factor 0.78. Start with the guess of 85 degree length. Adjust this value until you obtain a parallel equivalent R(p) = 50 ohms. This turns out to be 86 degrees, which at 21 MHz and vf 0.78 equals 8.73 feet. Thus the stub is inserted 8.73 feet from the antenna. The stub (or discrete component) must provide +34.44 ohms (inductive) reactance in order to cancel the - 34 ohms parallel equivalent (capacitive) reactance at the stub insertion point. This can be accomplished with a 3.74 foot shorted stub as shown above.



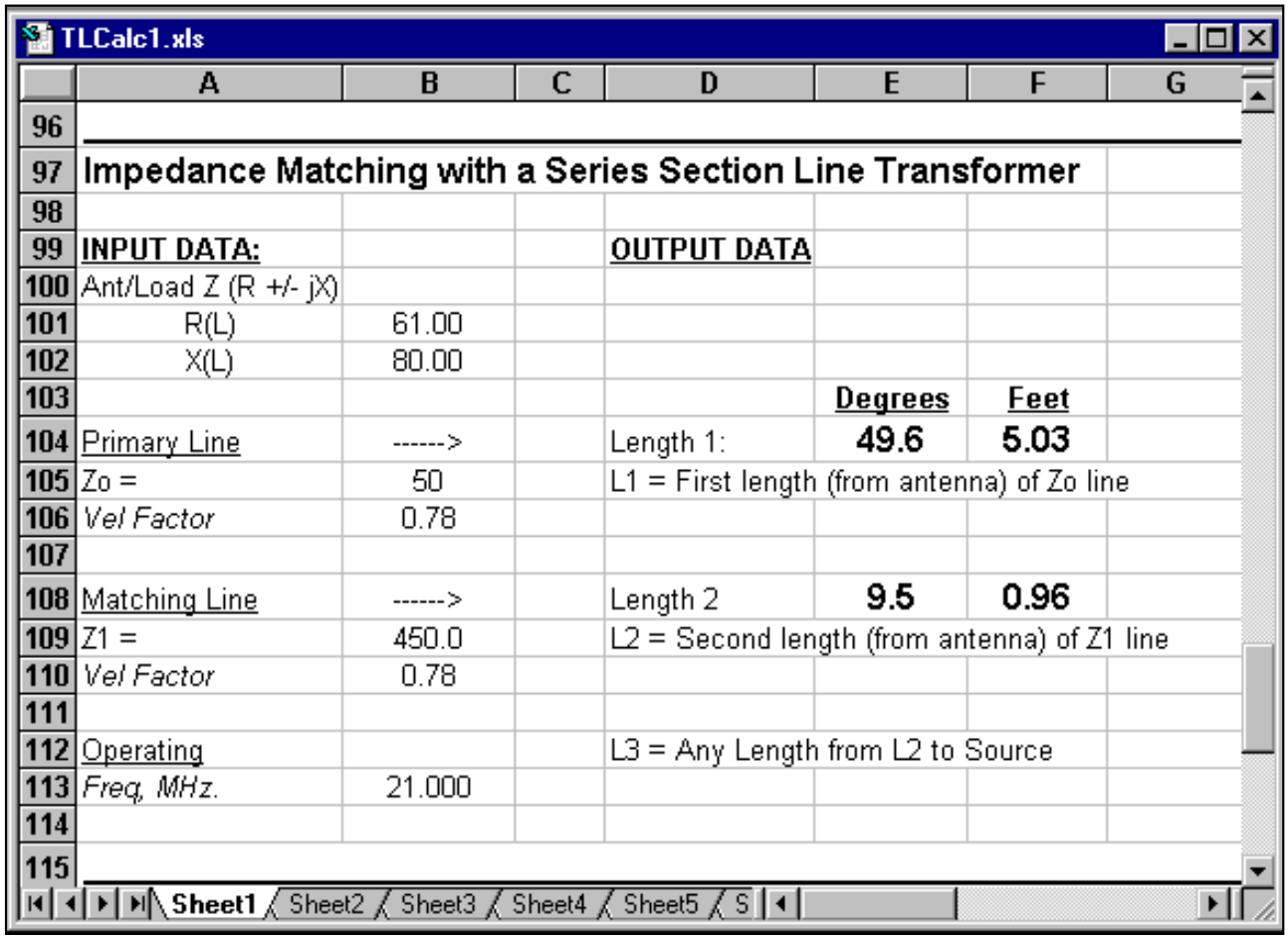
TLCalc1.xls

	A	B	C	D	E	F	G	H
52								
53	**Impedance (Z) at any Single Point along a Lossless Transmission Line:**							
54								
55	**INPUT DATA:**			**OUTPUT DATA**				
56	Ant/Load Z (R +/- jX)			Line/Input Z (R +/- jX)		Parallel Equivalent Z		
57	R(L)	61.00	------>	R(in)	**16.0**	R(in/p)	**50.72**	
58	X(L)	80.00		X(in)	**-23.6**	X(in/p)	**-34.44**	
59	Length of Line (deg)	86						
60	Freq (MHz.):	21		SWR:	**3.88**			
61	Vel Factor:	0.78		Length (Feet):	**8.73**			
62	Line Zo (R only)	50						
63								
64	* ***Caution:*** *Valid only for lossless lines. For lossy lines (high SWR or long lengths), validate outpu*							
65								

Sheet1 / Sheet2 / Sheet3 / Sheet4 / Sheet5 / Sheet6

Series Section Line Transformer (analytic solution) for impedance matching

While the theory of Series Section Line Transformer matching is beyond the scope of this article, the solution provided is analytic and straightforward. The same example used above for stub matching illustrates how it may be performed using a SSLT. In this case we elect to use 450 ohm ladder line with a vf = 0.95 for the matching section. The output indicates that a 0.96 foot length of ladder line should be inserted 5.03 feet from the antenna.



TLCalc1.xls

	A	B	C	D	E	F	G
96							
97	**Impedance Matching with a Series Section Line Transformer**						
98							
99	**INPUT DATA:**			**OUTPUT DATA**			
100	Ant/Load Z (R +/- jX)						
101	R(L)	61.00					
102	X(L)	80.00					
103					**Degrees**	**Feet**	
104	Primary Line	------>		Length 1:	**49.6**	**5.03**	
105	Zo =	50		L1 = First length (from antenna) of Zo line			
106	*Vel Factor*	0.78					
107							
108	Matching Line	------>		Length 2	**9.5**	**0.96**	
109	Z1 =	450.0		L2 = Second length (from antenna) of Z1 line			
110	*Vel Factor*	0.78					
111							
112	Operating			L3 = Any Length from L2 to Source			
113	*Freq, MHz.*	21.000					
114							
115							

Sheet1 / Sheet2 / Sheet3 / Sheet4 / Sheet5 / S

Download:

The spreadsheet was created in Microsoft Excel 7.0 format, with the filename TLCalc1.xls. It has been zipped for storage and downloading. If it is updated in any significant way, the next version will be named "TLCalc2".

- **Click here to download the AA3RL Transmission Line Calculator.**

(TLCalc1.xls). Excel 7.0 spreadsheet format, (~10K zip file).

This page is published by Mike Banz, AA3RL as a service to the Amateur Radio community.
Please distribute freely.

The author welcomes any questions, criticisms, or compliments via email.

email Mike with comments or questions.

Setting up shop

Some of the interesting shots of the W5AJ Rohn 25G tower

This installation is taking place in 2003.

And the gasoline driven digger was low cost compared to what it would have cost in time to recover.....

This is shot of project start.

Gasoline powered. Less pain then posthole diggers in WTX soil

AJ plate on Left Rohn plate is on the right

just south of the Crank up tower (see tower #1 page) Tower sections moved to location.

Hinged base. 1st ten feet in place

Concrete on base & guy posts was hand mixed 5000psi premix from hardware store. the 5000# version would setup to something like 2400psi strength in 3 days.

After getting ten feet up the section is laid over and another 20 feet added.

Setting the lower guys.

Bigger saw horse being put into place

Add another 30 feet & now 60 feet done - is ten open ?

note date on picture (I think 26th was highest SFI for the eleven year cycle)

Assembly of A4 with 40 meter kit.

that is a WTX Apple tree behind the A4

From the hinged end

20 feet of 25G on ground to right.

Note antenna assembly platform (some think this a trampoline)

Early Saturday morning SS SSB 2003

With the help of my wife the 20 foot tower for lift cable is raised

and everything is in place for a lift shortly after 8 am

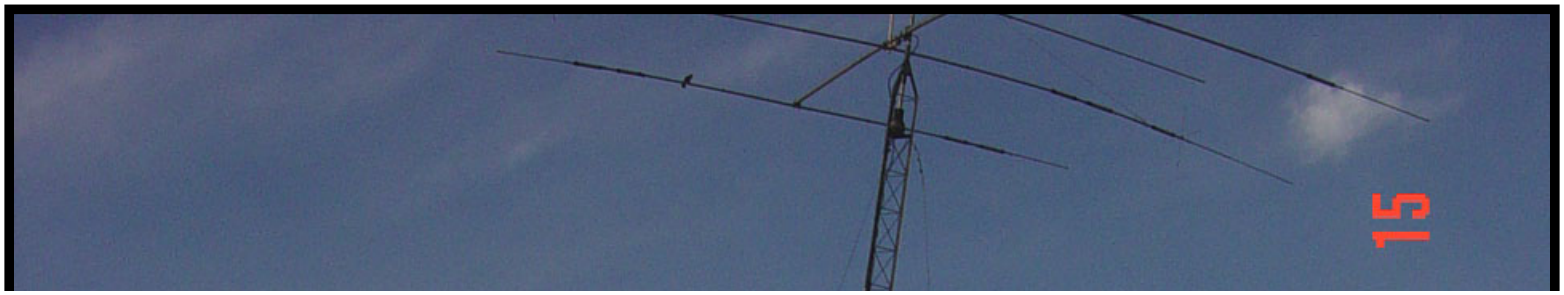
had planned this to be one person setup but had a guywire problem on the 20 foot tower.

tnx to the wife for bailing me out on that one.

there's another story there about always carrying a cell phone ....

The lift begins as the tower comes off the sawhorse

Stopped at this point to double check lift cable



Almost vertical - note bird already checking out the A4

The manual (no computer) calculations showed at this point the tower could be hand pulled

vertical

Wasn't a cake walk either

Taken just before sunrise Monday morning after SS SSB weekend

small print - this is not a design page - this shows the installation of a rohn 25G using a hinge plate similar to Rohn's. Hours of calculations were done to match conditions, tower & equipment and insure nothing was over stressed.

Work going on inside the shack building upper shelf

==========

in the background sound you are listening to a QSO with cv5h

that is used in the UBN study by W5AJ

W5AJ HOME PAGE

TOP 50

This page added 2003 Nov & is developing as time permits note not all pictures and text maybe uploaded

---

This page maintained by *W5AJ*

Please send your comments and suggestions to rwood90 #on# direcway.com

104060

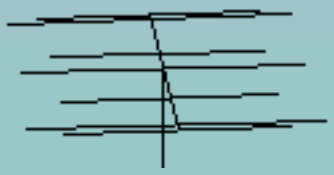
ServuStats

NYLO'S NOTEPAD

Homebrew projects for radio & electronics enthusiasts

- TOWER PAGE -

back to HOME page



There is a huge amount of relevant information you should know before building a tower. Here's a collection of information that I have gathered that will give you a boost. You'll be glad you took the time:

N1LO's GUYED TOWER TOPIC SUMMARY

To find out more about how towers react to windloads and why it is so important to select guy materials and sizes for low stretch as well as adequate ultimate breaking strength:

Guyed Tower Safety Analyses by Kurt, K7NV

OK, that's a lot of heavy reading so far.... Check out some of these photos and illustrations:

N1LO'S TOWER PICTURE GALLERY

PAGE 1

PAGE 2

PAGE 3

Have you found something useful here? Got a comment? Drop me a line: n1lo (remove 'nospam' from address when sending)

You are visitor number **since 12/21/99**

KO4BB's Tower Building Project

Last Updated:

NOTE: This is a description of how I built my tower. It is provided in the hope that it will be useful but without any warranty of any kind. Make sure to get the necessary permits for your area, if applicable and if you have any doubt, talk to a professional civil engineer even if not required by local regulations.

Even though a permit was not required in my county and subdivision (not required for ham towers under 75'), I thought about getting one for the piece of mind. However, in my county, the tower and antennas would have to be rated 140 mph to get a permit, something few ham radio budgets can afford, and something unnecessary in my case since the tower folds. Further, I had access to persons knowledgeable about building codes and permits and the final design takes into consideration the advice I got informally from these qualified individuals. While this is not as good as "money in the bank", or a building plan certified by a professional engineer, I felt is was sufficient for my purpose. Your mileage may vary.

If you intend on using information on these pages to build a permanent (non-folding) tower, make sure you understand the differences and protect yourself accordingly, both from safety and legal stanpoints.

Pictures: taken during construction, the finished tower and antennas.

Useful tower building resources:

- N1LO's home page contains a tremendous amount of information on tower building and anything related.
- K7NV's page on wind load calculations (this page has not proven to be reliably available, so you may want to check the links below).

Here is a link to his YagiStress software (download the demo version).

- WB6WUW's very informative page, which was featured in July 2003 QST.

This page would not be complete without the deepest thanks and appreciation for the members of the Playground Amateur Radio Club PARC, W4ZBB who came in droves to help with the tower raising party.

My Tower Project

I bought this folding Rohn 25 tower used from a local ham who had it attached to the side of his house. For extra safety (we are on the Florida coast, and we do have hurricanes and strong winds), he also had installed guy wires.

The guy wires were going from just below the hinge to the top of 7' high, 3" diameter schedule 40 steel pipes instead of directly to the ground because there was not much room where his tower was located and the guy wires would have been too steep to be effective.

I thought this would be a neat idea to reduce the risk of someone (like a kid) running into the guy wires. With the normal mounting scheme, guy wires leave the ground around 45 degree angle and that means that shrubs or a fence have to be installed to prevent accidents, and I certainly did not want to loose that much real estate. I also wanted to put the tower as close to the fence as possible. With the guy posts, I loose much less real estate and the guy wires take off from the tower closer to horizontal by about 15 degrees, so they should do a good job. The guy posts are at about 25 feet from the base of the tower.

So I did the same thing, except that in my case, the tower is detached from the house and squarely in the middle of my backyard.

To stiffen the guy posts, another EHS guy cable runs from the top of the guy post to a J bolt about 1.5' from the base of the post. This does little for ultimate strength but does stiffen the pipes and reduces flexing in the wind.

The pipes are also filled with concrete all the way to the top to make sure they do not collapse under extreme stress. The concrete makes them stronger and prevents moisture and water from getting inside the pipe and causing corrosion. The concrete also dampens the tubes, so they don't sing like a bell when something hits them.

Because it is a folding tower (it folds approximately in the middle, a design no longer offered by Rohn unfortunately), there are 4 guy wires instead of the usual 3 so that the fulcrum that allows to fold and raise up the tower has clearance.

Note: I read that Rohn stopped making the folding type tower because hams were overloading them and having problems. The tower is certainly very vulnerable when it is being cranked up and down, particularly when it is at a 90 degree angle. I wanted to take it down during moderate winds at one time, and the effect of the wind on the structure when the tower is folded is significant. You do not want to leave it like that very long. However, when the tower is straight up and properly rigged, it is as strong as a normal Rohn 25 tower.

The guy posts are actually 12' long 3" diameter schedule 40 steel pipes (structural grade, not just "pipe") with 5' in the ground, in concrete. The guy post holes are therefore 5'6" deep, and 1.5' x 2.5' and filled with 3000 lbs rated strength, fiberglass reinforced concrete. The post base also holds the J bolt for the post guy. Place a brick at the bottom of the hole so the post is at least 3" off the dirt. I poured concrete up to about 3" under the ground surface, so that if and when I move out of the house, I just have to cut the pipes flush with the concrete and cover the hole with dirt and grass.

The tower concrete base also is filled with 3000 lbs rated, fiberglass reinforced concrete. I selected the fiberglass reinforcement because the tower base is bolted down in the concrete using 4 J bolts, instead of just burying a tower element as I have seen it done often. The fiberglass gives the concrete an extra measure of strength around the J bolts and only adds about $5.00 per cubic yard to the cost of the concrete. In total, I used 5 cubic yards of concrete, in addition to what I used to fill the posts (I used regular Quickrette for that). 5 yards was the minimum quantity for which there was no delivery charge from all the vendors in the area.

Please note that fiberglass reinforcement does not allow to get rid of the rebar cage unfortunately.

To make sure the concrete hardens and does not just dry, keep it covered from the sun and keep it moist. I used an old piece of carpet laid over the concrete pad, and covered with a large piece of cardboard. I waited until the next day in the morning, when the concrete surface was already somewhat hard. Then twice a day I would hose the carpet and cover it with the cardboard so it did not dry too quickly. I took carpet and cardboard off after about 4 days of that regiment. The concrete looks pretty hard to me. I pounded on it with a hammer and could not chip it.

I let the concrete cure for two weeks before putting up the tower. The concrete was plenty hard enough by then and we had absolutely no problem, even though it was

interesting when two of us were hooked up to the tower at about 20' up in the air before we had the guy wires attached (please note we had rope guys for safety while we were doing that, but these don't do much to stabilize the tower to the point I felt I needed for peace of mind.)

Go to pictures
Return to KO4BB's Ham Radio Page.

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Last Updated:

Amateur Radio

N9XT, I am a Life member of the ARRL, I have been licensed since 1977 and have worked 190 Countries (I just worked 3D2CY Conway Reef) 1st DX contact in 18 years. My other Passion is UHF and Control Systems.

Life Member

The TH7DX antenna is up! Scroll down to see pictures. See you on the airways!

Updated 12/25/2003

Anyone with a Website, particularly if you have projects or DX Pedition photos that would like a link please contact me.

The American Radio Relay League

Celebrity Hams

- Home
- Nascar
- Desert Fun
- Amateur Radio
- Celebrity Hams
- My BBQ
- Genealogy
- Recipes n Tips
- Science_Invention
- Beverages

DRONK RADIO NETWORK

I took this picture from Mt Soledad in San Diego in November of 1971. The strike hit a sub station and blacked out an entire area of town.

AC6V has the most active Links

ARRL

Radio Anomalies

QRZ

Extra Class Question Pool

Radio Frequency Spectrum

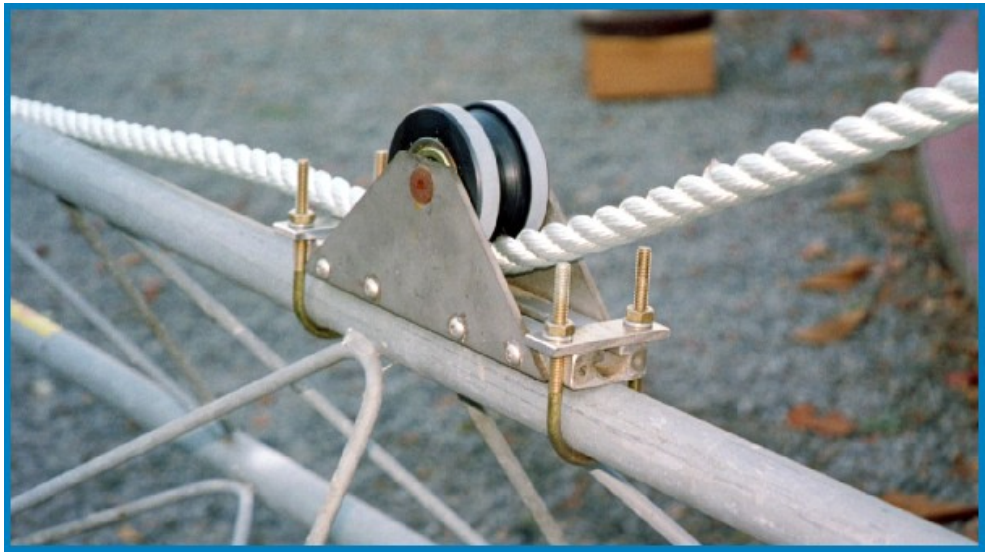
W4RNL

101 Science

Many Local Links, courtesy of AC6V
Find a store, Amateur activities, Repeater lists and more!

You need a level, elevated, surface to assemble this antenna. This is how I did mine.

I used a Carpenters bubble level with a riser block on each side to verify that all elements were in plane. (Level not shown)



I made this pulley to support the mid section
while raising the assembly.

The top of the tower is fitted with this larger pulley.

You can see how this works. There is a double pulley (2 grooves side by side) on the roof.

The other end of the ropes is attached to the Blazer. The rope is a loop as you can see by the previous photo.

We used an extension ladder at the right height to support the tower and keep the antenna off the ground. The ladder is lashed to the tower and follows the tower as it is raised to this point.

Art, K6XT does all the finishing touches to the assembly, here he is RTVing the Balun to weather proof it.

We are read to raise the Tower assembly, we think!

The Assembly is about 45 degrees up at this point.

The Assembly is almost vertical now. Damon is getting ready to nest the tower into a roof eave bracket we made to hold it in place.

First contact 3D2CY Conway Reef!

Coming soon, an interactive calculator to determine the force required to upright your tower and antenna. I'm looking for assistance to do this.

This is intended as a guide only, The Web Master (N9XT) is not responsible for accidents that may occur through miscalculations, faulty gear or poor procedures.

This is it, The vintage is mid 70's Drake and Alpha. Some day we may upgrade to the state of the art stuff. But you know what, they answer every time I call so what's the point? N9XT

N9XT has more at this site!

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SPECIAL ANTENNAS

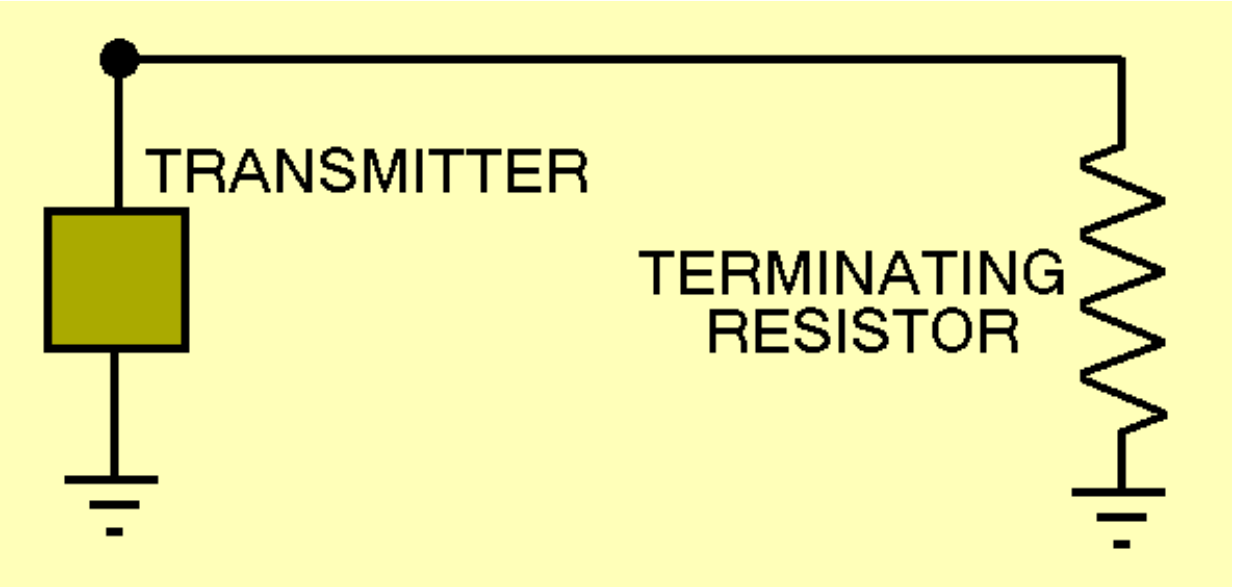
In this section we will cover some special communications and radar antennas. Some of these antennas we touch on briefly since they are covered thoroughly in other courses.

Previously discussed antennas operate with standing waves of current and voltage along the wires. This section deals principally with antenna systems in which the current is practically uniform in all parts of the antenna. In its basic form, such an antenna consists of a single wire grounded at the far end through a resistor. The resistor has a value equal to the characteristic impedance of the antenna. This termination, just as in the case of an ordinary transmission line, eliminates standing waves. The current, therefore, decreases uniformly along the wire as the terminated end is approached. This decrease is caused by the loss of energy through radiation. The energy remaining at the end of the antenna is dissipated in the terminating resistor. For such an antenna to be a good radiator, its length must be fairly long. Also, the wire must not be too close to the ground. The return path through the ground will cause cancellation of the radiation. If the wire is sufficiently long, it will be practically nonresonant over a wide range of operating frequencies.

LONG-WIRE ANTENNA

A LONG-WIRE ANTENNA is an antenna that is a wavelength or longer at the operating frequency. In general, the gain achieved with long-wire antennas is not as great as the gain obtained from the multielement arrays studied in the previous section. But the long-wire antenna has advantages of its own. The construction of long-wire antennas is simple, both electrically and mechanically, with no particularly critical dimensions or adjustments. The long-wire antenna will work well and give satisfactory gain and directivity over a frequency range up to twice the value for which it was cut. In addition, it will accept power and radiate it efficiently on any frequency for which its overall length is not less than approximately 1/2 wavelength. Another factor is that long-wire antennas have directional patterns that are sharp in both the horizontal and vertical planes. Also, they tend to concentrate the radiation at the low vertical angles. Another type of long-wire antenna is the BEVERAGE ANTENNA, also called a WAVE ANTENNA. It is a horizontal, long-wire antenna designed especially for the reception and transmission of low-frequency, vertically polarized ground waves. It consists of a single wire, two or more wavelengths long, supported 3 to 6 meters above the ground, and terminated in its characteristic impedance, as shown in figure 4-34.

Figure 4-34. - Beverage antenna.



Q.44 To radiate power efficiently, a long-wire antenna must have what minimum overall length? **Answer**

Q.45 What is another name for the Beverage antenna? **Answer**

V ANTENNA

A V ANTENNA is a bidirectional antenna used widely in military and commercial communications. It consists of two conductors arranged to form a V. Each conductor is fed with currents of opposite polarity.

The V is formed at such an angle that the main lobes reinforce along the line bisecting the V and make a very effective directional antenna (see figure 4-35). Connecting the two-wire feed line to the apex of the V and exciting the two sides of the V 180 degrees out of phase cause the lobes to add along the line of the bisector and to cancel in other directions, as shown in figure 4-36. The lobes are designated 1, 2, 3, and 4 on leg AA', and 5, 6, 7, and 8 on leg BB'. When the proper angle between AA' and BB' is chosen, lobes 1 and 4 have the same direction and combine with lobes 7 and 6, respectively. This combination of two major lobes from each leg results in the formation of two stronger lobes, which lie along an imaginary line bisecting the enclosed angle. Lobes 2, 3, 5, and 8 tend to cancel each other, as do the smaller lobes, which are approximately at right angles to the wire legs of the V. The resultant waveform pattern is shown at the right of the V antenna in figure 4-36.

Figure 4-35. - Basic V antenna.

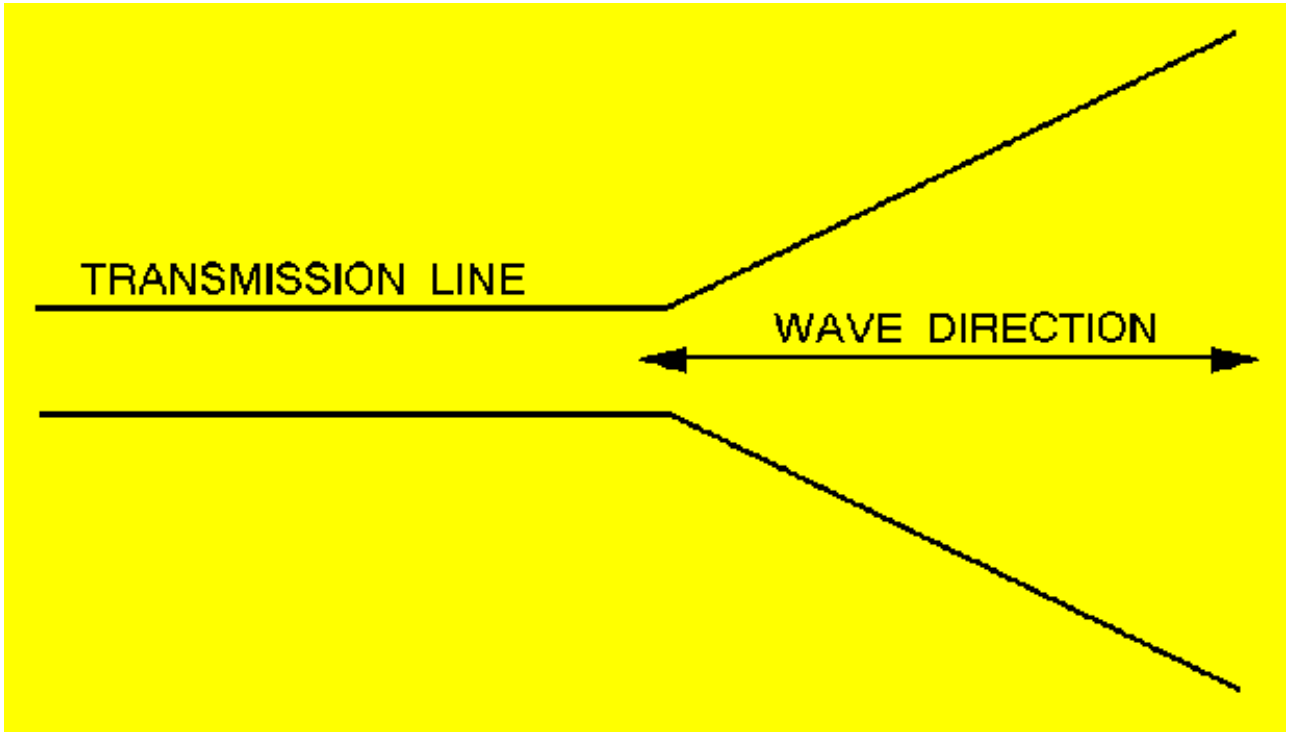
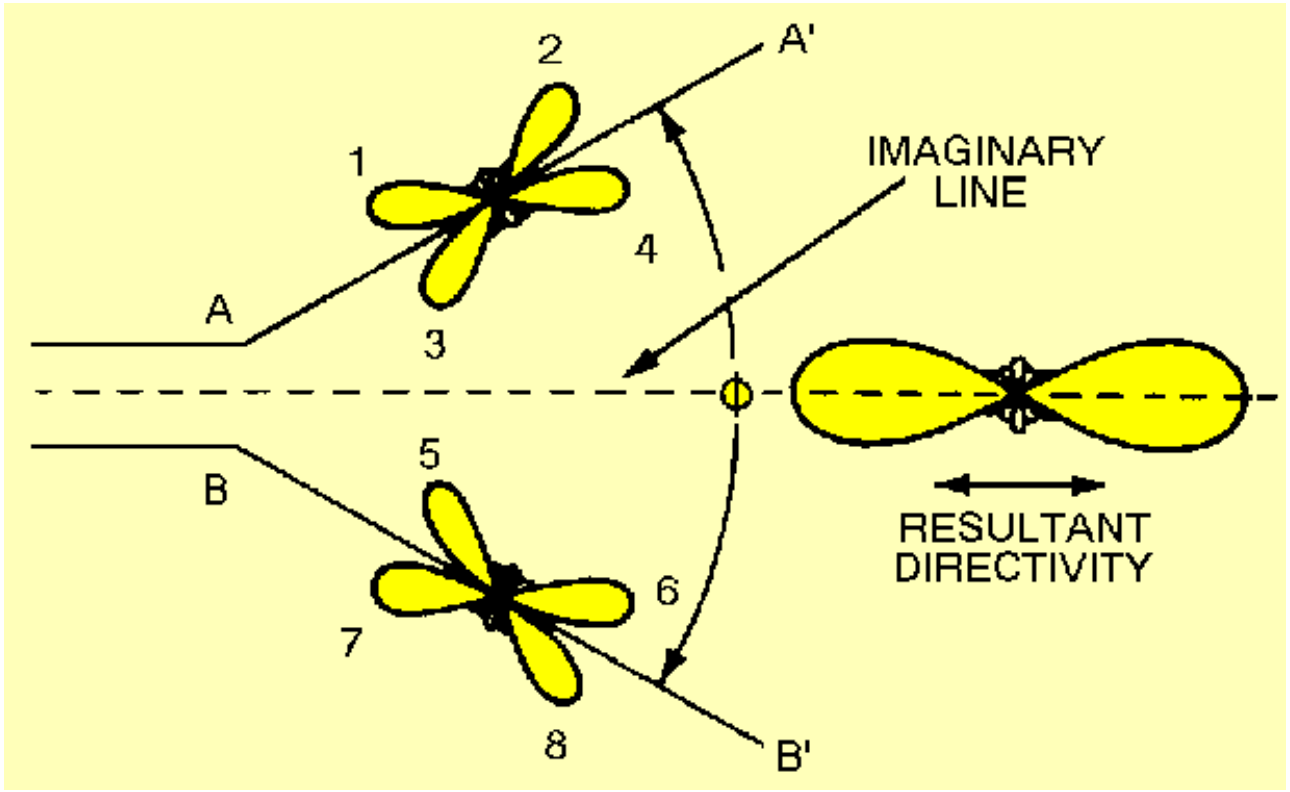


Figure 4-36. - Formation of directional radiation pattern from a resonant V antenna.



Q.46 What is the polarity of the currents that feed the V antenna? Answer

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The RXO Unitenna

by R. D. BROWN G3RXO
email: standardtrees@btconnect.com

WHILST touring New Zealand some years ago, I needed a portable all band HF antenna and conceived the following design concept which produced surprisingly good results. Back in the UK, I did have time to construct and test it and it worked surprisingly well. It is extremely cheap, extremely light, and works continuously over an extremely broad bandwidth. It is non-resonant, vertically polarised and out-performs a ground plane throughout a 3:1 frequency range. Accepting a loss of some 3 dB from a ground plane, its useable frequency range is some 5:1. It has very low angles of radiation throughout the range, requires no radials, and is not susceptible to static.

My thinking was along the following lines:

(1) Consider a vertical dipole, it's current distribution and resultant polar diagram.

(2) Consider it as a folded dipole.

(3) Consider it's current distribution and polar diagram if it is fed off centre. The dissimilar currents in the two elements of the dipole are within one evanescent induction field and therefore cumulative as to radiation which can give rise to an overall current distribution and polar diagram which changes very little with change of frequency.

(4) By feeding at approximately one third of the way along the dipole the current distribution and polar diagram vary approximately from that of a vertical dipole to that of a vertical quarter wave over a frequency range of some 3:1; for example from 7 MHz through to 21 MHz.

(5) Above the antenna's maximum useable frequency the polar diagram becomes split and of high angle.

(6) As the applied frequency is reduced below its normal range the polar diagram remains much as for a vertical quarter wave, but with diminishing efficiency, although still useful over a frequency range of some 5:1.

(7) Since the antenna is a closed loop it is less susceptible to static.

(8) It requires no radials and therefore incurs no ground losses.

(9) It is fed with a tuned line which should be as perpendicular to the antenna as possible.

(10) Variation of the feed point to about 25 per cent from the antenna's end results in slightly better current distribution, but reduced bandwidth. The optimum point is around 30–35 per cent from the end.

(11) It is cheap to produce, lightweight, portable, unobtrusive and has virtually no power limits.

The original antenna was made of regular single core conductor taped down each side of lengths of plastic waste pipe and guyed with bricklayer's line. For example; practical dimensions are a 9.2 m dipole, fed 3 m from base and about 1 m above ground; coverage 7 MHz to 21 MHz continuously with full efficiency. Pro-rata for other frequency ranges.

I found the antenna to be surprisingly quiet with respect to background noise and very effective at low angles and long range. I hope that others will give this design a try and would be most interested to learn of their results with it. βi

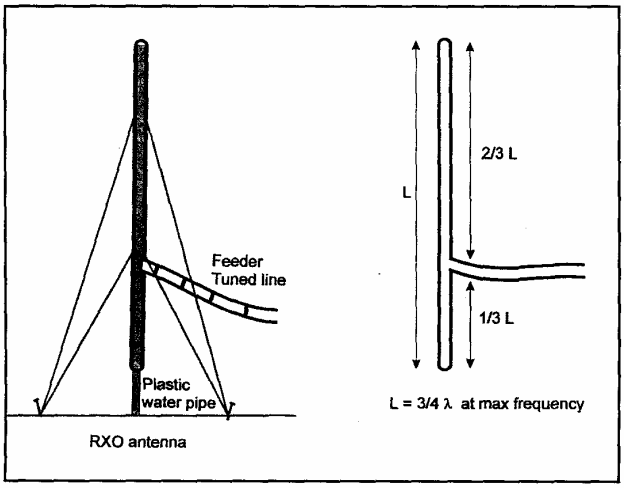

Figure 1.

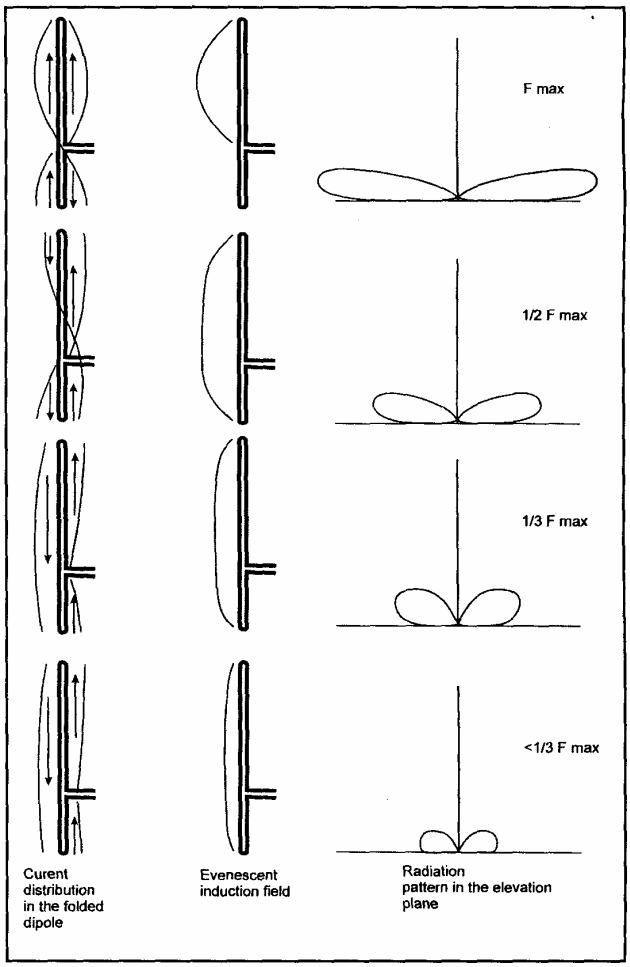

Figure 2.

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TRANSPORTABLE VERTICAL ANTENNA

This vertical was built for me by Luis, CT1BYR, a skilled ham and homebrew fan - everything from smd component level to large antenna array construction.

The vertical's slim silhouette stands almost 10 metres high.

After numerous conversations with Luis regarding transportable antennas he came up with this solution for me to take on my trips to the Azores. Knowing that I often use an automatic SGC-230 antenna coupler and so tuning would not be a problem, he suggested combining this with a random length vertical. Operating away from home is also about having redundancy plans - this antenna can also be used if the SG-230 is unavailable as a standard quarter wave vertical, on most of the HF bands, simply by altering its length.

The antenna stands over 9 metres high when fully erected and has four levels of guying cables to withstand the harsh weather associated with this part of the world. Built from stainless steel, instead of aluminium, this provides added weather resistance and durability. As does its slim silhouette for minimum wind resistance. Thin walled stainless steel piping was easier to find at our local hardware store and although slightly heavier than aluminium this was not a problem as it is still easily carried by one person. The base is made from a nylon block with a deep circular recess to accommodate the bottom section and isolate it from ground. A couple of bolts are also screwed to this same block to attach the antenna's RF ground wires - the more the better.

To allow transportable operation and easy set-up, Luis used a telescopic solution for the

CT1BYR preparing to raise the vertical.

vertical element. Five 2 metre pipes all slotting inside each other reminds us of a car radio antenna on a larger scale. Each of the lower four section ends have a guying clamp to affix three guy ropes and when tightened, give the vertical rigidity and prevent the slimmer elements from slipping back inside the larger ones. This gives us a total of twelve guy wires which may be reduced for calm weather operating. In fact, the bottom level of three guy ropes are practically only used to facilitate raising the remaining upper sections by a single person. It saves him having to hold the bottom section vertically steady while telescoping out and tightening the remaining four sections.

Details of the telescopic antenna sections with guying clamps.

Lowering the antenna is just as easy as raising. For storage or travel, the vertical element, ground wires and guy cables fit inside a 2 metre long cardboard or plastic tube making it easily transportable and an ideal antenna for emergency, field-day or dxpedition use.

Back to Fun Stuff

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During March-June 2004, visitors averaged 200 per day.

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These are DOS-executable self-contained programs each occupying a single binary file. No use is made of external software. This ensures portability between DOS/Windows machines.

There is a brief one-line description following each program's name.

Approximate file sizes are from 25 to 70 kilobytes. Download, or run without downloading, in a few seconds. Just click on a program's underlined name.

Download from here to a Windows Desktop icon or collect programs in a common Windows folder of your own chosen name. There's no un-zipping inconvenience. Programs can be run immediately after download.

To run a program click on its desktop icon. If the program does not entirely fill the screen then look along the top tool bar and click on the button which expands the display outwards.

The mouse is not available when the program is running full screen. Close the program from within by hitting either the 'E(xit)' or 'Q(uit)' key when displayed in the menu on bottom line 25. You will be immediately returned to the desktop.

If the program aborts into DOS due to entry of 'impossible' data, to return to Windows Desktop type the DOS command 'EXIT' against the C>prompt followed by the enter/return key. Or to run the program again type the program's name alongside the C>prompt. No harm will have been done to your computer, operating system or to the program itself.

Remember, when in DOS, the DOS 'EXIT' command, followed by the 'Enter' or 'Return' key, will always return you to the Windows DeskTop. No need to get lost!

RJELOOP1 *
Performance of Transceiving, single-turn, magloop antennas of various regular shapes.

RJELOOP2 *

Performance of Transceiving, single-turn, magloop antennas of rectangular shape.

RJELOOP3 *
Analysis of receiving, multi-turn, square, loop (or frame) antennas, ELF to HF.

HELICAL3 *
Continuously-loaded, helically-wound HF antennas, top-capacitance loaded with a rod.

TOROID *
Ferrite & iron-dust cored, toroidal coils. Dimensions, Core permeability, Turns, uH, pF, Freq.

VERTLOAD *
Base-fed vertical antennas, coil-loaded at any height, with coil design.

RADIOETH *
DC, power freq & RF characteristics of a single shallow-buried ground wire.

EARTHRES *
Ground electrodes. Resistance msmnts and predictions, rods, wires, plates, mats,

TANT136 *
LW & MW performance of small T-antennas above a system of buried ground radials.

ENDFEED *
HF performance of Inverted-L antennas above a system of buried ground radials.

LINEAR1 *
Models simple Class-AB, Push-pull, Bipolar Linear RF power amplifiers up to 30 MHz.

BALUN4 *
Design and performance of HF transmission line ferrite-core transformers, Z-ratio 4-to-1.

RJELINE1 *
Performance of openwire lines, Power freqs up to UHF, for any complex termination.

RJELINE2 *
Full analysis of balanced-twin lines, 20Hz-1GHz, for any complex termination.

COAXPAIR *
Full analysis of coaxial lines, Power freqs to UHF, for any complex termination.

RJELINE3 *
Full analysis of balanced lines with facilities for HF line xfmr design.

TRAPDIP *
Design and performance of 2-band trapped dipole antennas including trap design.

PHASENET *
Design of T & Pi phase-shifting networks for use with antenna arrays.

TPINET *
Design of T & Pi impedance-matching and phase-shifting networks.

SWR-FREQ *
Compare simple & folded dipoles. Input Z. Feedline VSWR vs freq. Bandwidth.

SOILSKIN *
Enter soil R and K characteristics. Display a table of skin depth vs frequency.

COAXRATE *
Power ratings of solid polythene coax cables vs dimensions and SWR. 0.3 to 3000 MHz.

ZL_ZIN *
Compute antenna feedpoint impedance from measured line input impedance, known Zo & length.

SOLNOID3 *
Single-layer, air-core coils of all proportions. Dimensions. Turns. Temp rise vs applied volts.

MAGLOOP4 *
Performance of regular-shaped magloop antennas versus height and type of ground.

PADMATCH *
T, Pi and L resistive matching-pad design using standard-range resistor values.

MIDLOAD *
Design of very short, centre-loaded dipoles above lossy ground. With coil design.

LCR *
Full analysis versus frequency of behaviour of an L+C+R parallel-tuned circuit.

SWRMETER *
Modelling, design, calibration & operation of HF SWR meters using a ferrite current xfmr.

LOADCOIL *
Design of short vertical antenna + loading coil. Slide coil up/down to maximise effncy.

PI_TANK *
Design Pi-match output cct for RF power amp. Effncy, harmonic suppression, phase shift.

ZOC_ZSC *
Calculate Trans-line Zo,L,C,R,G,dB,VF from open and closed input impedance measurements.

Pi_L_Net *
Design Pi and Pi-L output networks for RF Power amps. Effncy, harmonics, phase delay.

TETRODE1 *
Design of Class A, AB1, AB2 and C RF Power Amplifiers, using beam tetrodes or pentodes.

TOPHAT2 *
Short vertical antenna with Top-capacity-hat made with N radial wires. L-tuner values.

RJELINE4 *
Analysis of balanced-pair transmission lines. Line xfmrs with complex terminations.

T_TUNER *
Design & power efficiency, etc., of T-match tuning networks for transmitting antennas.

GRNDWAV3 *
Groundwave propagation. Field strength vs pathlength, terrain and frequency, VLF to HF.

DIFFRAC1 *
Radio propagation. Diffraction over a single obstruction. Tx output to Rx input.

DIFFRAC2 *
Radio propagation. Diffraction over two successive obstructions. Tx output to Rx input.

LINOSITE *
Simple Line-of-sight radio links. Approximate aperture, beamwidth, propagation loss etc.

MATCHSEC *
Impedance matching by inserting in antenna feedline a short section of different Zo.

BOTLOAD2 *
Inductance and bottom-loading coil design, or tune pF, for vertical or sloping antennas.

L_TUNER *
Calculate coil & capacitor values of an L-Network to match any pair of complex impedances.

WIRESKIN *
Skin resistance vs frequency of a single wire, an open-wire line and a dipole.

SELECT_1 *

Simple receiver preselector is a tapped parallel tuned circuit. Also coil design.

TWOCOILS *
Design & position two common-axis solenoids for desired coupling and mutual inductance.

DIPOLE3 *
A dipole at any height + balanced line + balun + coax line + L-match tuner L & C values.

DIPCAGE2 *
Cage dipole of N wires, resonant length, bandwidth vs SWR, end-effect, feedpoint impedance.

ADDALOAD *
Location and value of L or C loading component needed to resonate an antenna wire.

TRIODE1 *
Triode RF power amplifiers. Class-A, AB, C. Tuned tank or Pi-match output cct component values.

MULTILAY *
Design of multilayer coils wound in bobbins or other formers. 3 MHz and below.

WINDOM2 *
Performance of 1/2-wave horizontal dipole fed off-centre via vertical single-wire feedline.

FEEDPOWR *
Crude estimate of power radiated from coaxial line feeding 1/2-wave dipole without a balun.

COAXTRAP *
Design of antenna trap using length of coaxial line wound as solenoid on a coil former.

TWINTRAP *
Design of an antenna trap using length of twin-line wound on coil former. (Experimental)

OSCTRACK *
Superhet receivers. Optimum tracking of RF and oscillator tuned circuits. L & C values.

STUMATCH *
Match antenna input impedance to feedline with stub line xfmr. Balanced or coaxial lines.

STUBTUNE *
Match antenna input impedance to twin feeder with stub line xfmr. Includes xfmr efficiency.

LPF_HPF *
Simple L and C High and Low-pass filters. T and Pi Sections. Insertion loss vs freq.

BANDPAS1 *
Simple L and C bandpass filters. T and Pi Sections. Insertion loss vs freq.

INV_VEE *
Performance of resonant 1/2-wave Inverted-Vee and Dipole antennas including feedlines.

RHO_SWR *
Exact values of Rho and SWR on mismatched lines. Compare with measured values.

WHIP_1 *
Signal strength received on whip antenna + un-un transformer + coax line + receiver.

TWOHOPS *
Skywave Trigonometry and insight into how losses are proportioned along a radio path.

HARM_GEN *
Simple single-transistor harmonic generator. Performance and behaviour vs frequency.

R_X_SWR *
Feedpoint Impedance and SWR of an antenna in the vicinity of resonance.

CHOKEBAL *
Design and performance of a choke balun in conjunction with feedline and antenna.

BALCHOKE *
Better, recommended, version of program ChokeBal above. Changes in input data.

XTL_SET *
Design, performance, of crystal set, or preselector, using two coupled tuned circuits. From Tx to phones.

RADIALS2 *
Performance of a set of shallow-buried ground radials considered as transmission lines.

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Short Vertical Antennas and Ground Systems - VK1BRH

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Introduction

There have been a number of articles discussing the merits or otherwise of various types of ground systems. The analysis of such systems is complicated by the facts that practical ground system parameters are difficult to quantify, antennas are situated in less than ideal locations, and literature, that may provide insight into what is happening, often does not present the information in a practical or applicable form. Often the reader is left to extrapolate beyond the bounds of presented data and arrive at the incorrect conclusion. Some folk-lore has been generated as a result of these types of difficulty; one such lore is 'the higher the better', which is an over-generalisation if said without qualification.

Activities

I am interested in 160m operation; it is a challenge to develop reasonably efficient antennas for this*'top band'.* I am also interested in mobile and portable work so I constructed antennas and experimented a bit. It was difficult to quantify the results and performing alterations was rather tedious, so I turned to computer simulation.

After performing several different simulations I developed the feeling that it should be possible to optimise the efficiency of an antenna by altering its earth currents and through the interactions with the mutual impedance of the antenna and its ground image. This turned out to be a very fruitful study, although I initially thought the results were dubious!

Simulations

Fortunately non-ideal antennas have been under analysis by various organisations and many papers have been written covering the theoretical aspects. One organisation of note is the Lawrence Livermore National Laboratory (LLNL), which was commissioned by the U.S. Department of Defence to perform theoretical and practical analysis of various antenna systems. Some of this material has been declassified and is now publicly available.

During this period several computer-based antenna simulation programs were developed; of note is the Numeric Electromagnetic Code (MINI-NEC, NEC-2 and NEC-3 and now NEC-4).

NEC-81 is the name of the earlier PC version of NEC-2. NEC-81, NEC-2 and MINI-NEC are available through the Applied Computation Electromagnetic Society (ACES); contact Dr R.W. Adler, ECE Department, Code EC/AB, Naval Post Graduate School, 833 Dyer Rd, Rm. 437, Monterey California, 93943-5121, U.S.A or Fax 1 408 649 0300 or E-mail 554-1304@mcimail.com, for the conditions, membership and handling fees if you want to obtain these programs.

Post publication note: public domain software now available at Unofficial NEC2 pages.

NEC-2 can model antennas in proximity to lossy ground. The lossy ground analysis is based on work by Sommerfeld; and appears to yield realistic feed-point impedances and reasonably quantifiable losses.

System Performance

The simulation goal was to determine the relative merits of elevated groundplanes for short vertical systems, ie up to 0.25 wavelengths.

Traditionally, it has been convenient to measure the feed-point impedance of the lossy antenna system and compare it with the ideal (theoretical) case. The feedpoint resistance for vertical antennas is called the base resistance (Rb); Rb is composed of the useful radiation resistance (Rr) and the collective loss resistance (Rl). Rr can be evaluated theoretically, it would be the Rb of an antenna over ideal infinite ground.

Rr can be obtained by the additional simulation of the ideal ground model, effectively doubling the simulation time to obtain results. Fortunately, NEC-2 calculates the amount of power radiated around the region of the antenna and divides this by the applied power. This ratio depends upon whether the model is simulated in free space or over ground.

The radiation region of an antenna in free-space is a solid sphere so this ratio should be unity. The radiation region is a hemi-sphere when the antenna is situated against an infinite ground, in which case the value should be close to two; this is due to the ground reflecting power into the upper hemisphere (effectively up to 3dB of gain). This ratio is a also used to gauge the stability of the model; if its value greatly exceeds the expected value then the model has failed.

Post publication note: an antenna may exhibit up to 6dB of gain over its free-space radiation pattern due to ground refelection in some directions.

It is a simple means to use this power ratio to determine the antenna's efficiency. Note that the radiation resistance can be dervied via: Rr = Rb / efficiency; I used this analysis to cross-check the results of some simulations.

Results

The term displacement is employed for the height of the groundplane above ground; this avoids confusion with antenna element lengths. All lengths, heights and displacements are measured in wavelengths (lamda) unless otherwise indicated.

Resistance versus Groundplane Displacement

Figure 1 shows the effects of various displacements upon a 0.10 wavelength vertical antenna with three 0.10 wavelength radials at 1.825MHz. I chose the average ground parameters: relative dielectric constant 13 and conductivity of 5 milli-Siemens per meter (13,5); which are typical for dry clay and indicative of the Canberra region.

Note that the Rb is high at zero displacement, much higher than Rr, so Rb is largely composed of loss resistance at this point, and the antenna is primarily heating the ground! Notice how difficult it is to determine the optimal displacement from the Rb and Rr curves on this graph.

Efficiency versus Displacement

Figure 2 shows the antenna's efficiency over the same range. Notice that there is an initial peak at fairly low displacements. Advantage will be taken of this effect.

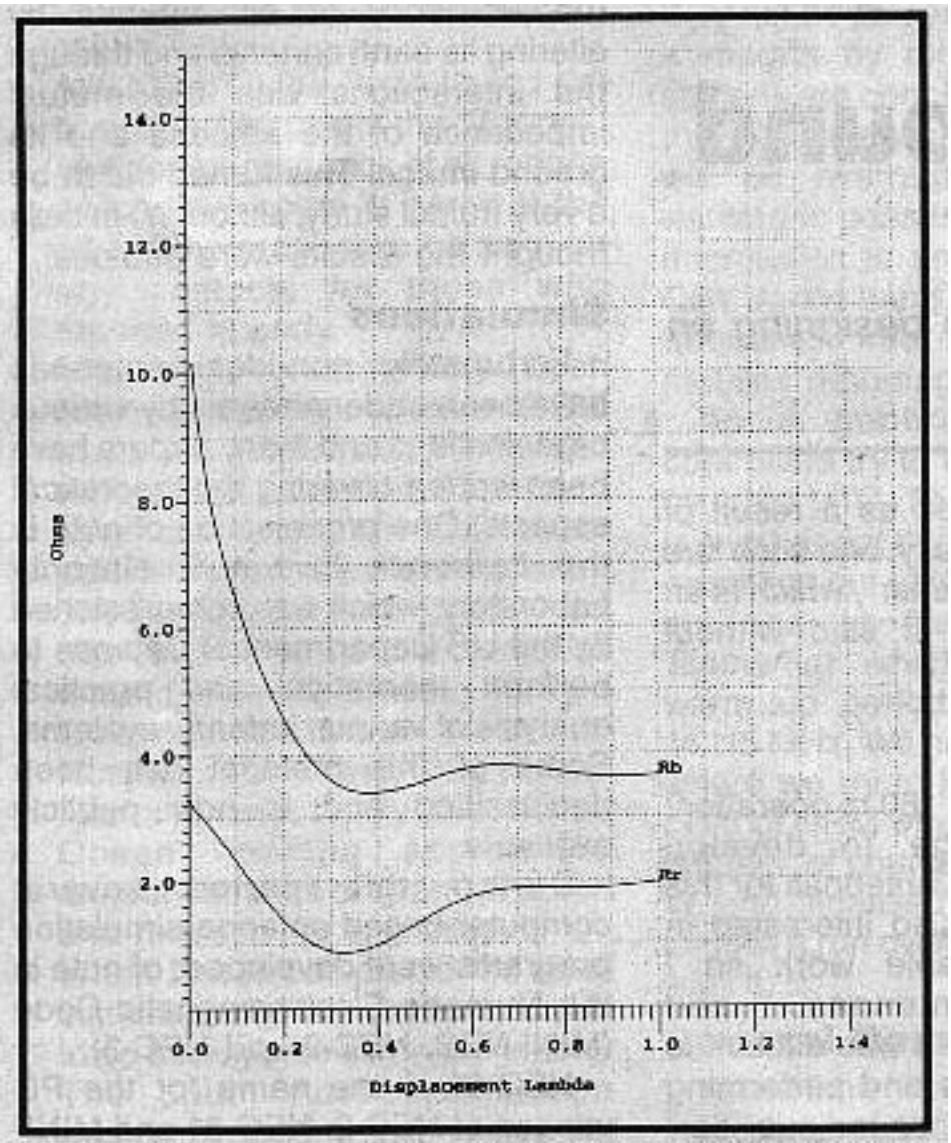


Fig 1 — 0.10 wave radiator/radials ` 3 @ 1.825 MHz (13,5).

Fig 1 - Resistance vs Displacement

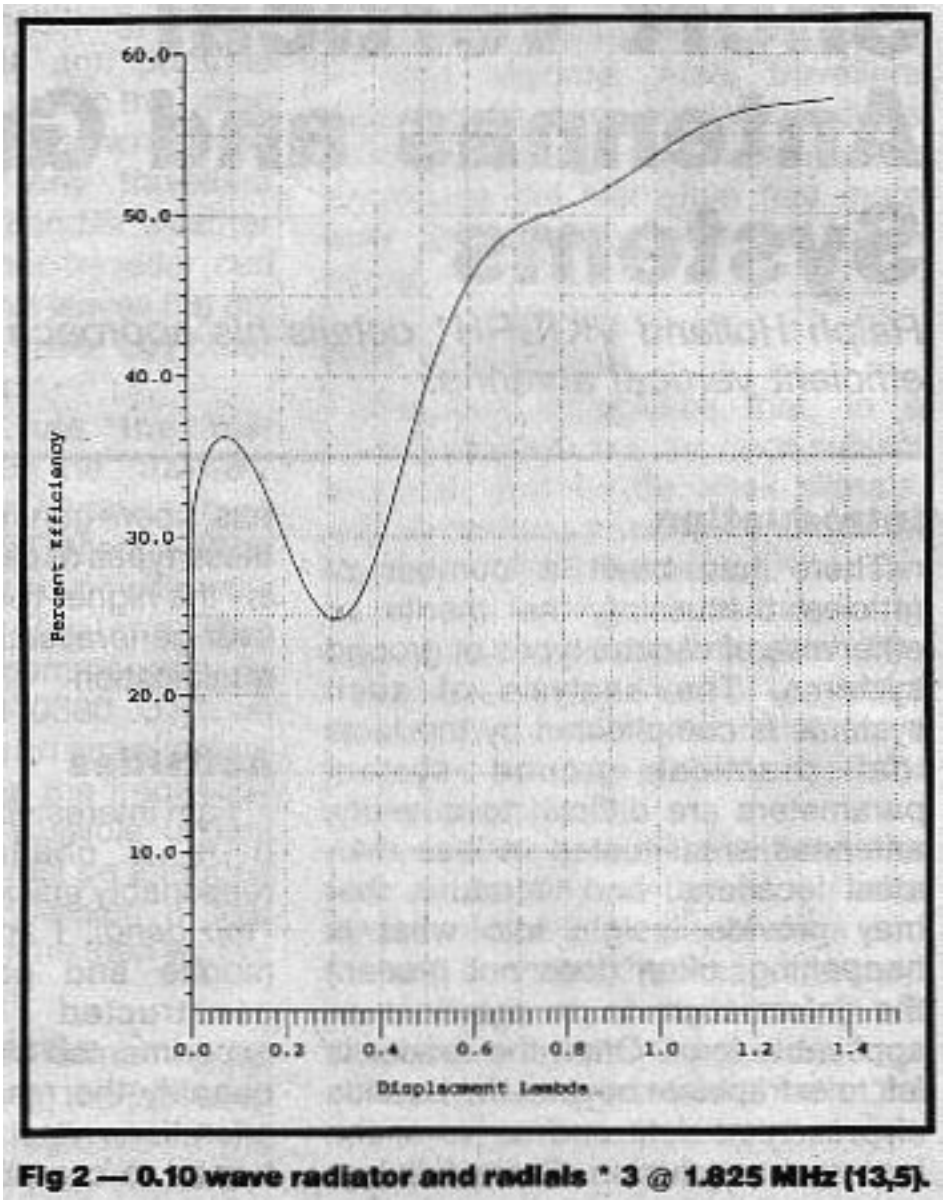


Fig 2 - efficiency versus displacement

Displaced Antennas versus the Number of Radials

Figure 3 is the comparison between the conventional groundplane, situated on or in the ground, and the elevated groundplane. The simulation parameters are 1.825Mhz, 0.25 wave radiator and radials, and ground parameters (13,5). The integer simulation points are joined to form curves (I did not actually simulate a non-integral number of radials).

Notice that the elevated curves are asymptotic to a horizontal line about 37 percent efficiency. The elevated curves are more than satisfactory at the three to four radial mark; the traditional ground screen doesn't even get near this at 32 radials. Recall that the typically quoted commercial design figure is to strive for 120 ground-based radials.

Variable Radiator, Radial Length and Displacement versus Number

of Radials

Figure 4 illustrates the effect of radiator and radial lengths. The simulation is at 1.82MHz and for ground parameters (13,5). The top curves represent the elevated groundplane, while the bottom curves are for ground-based antennas.

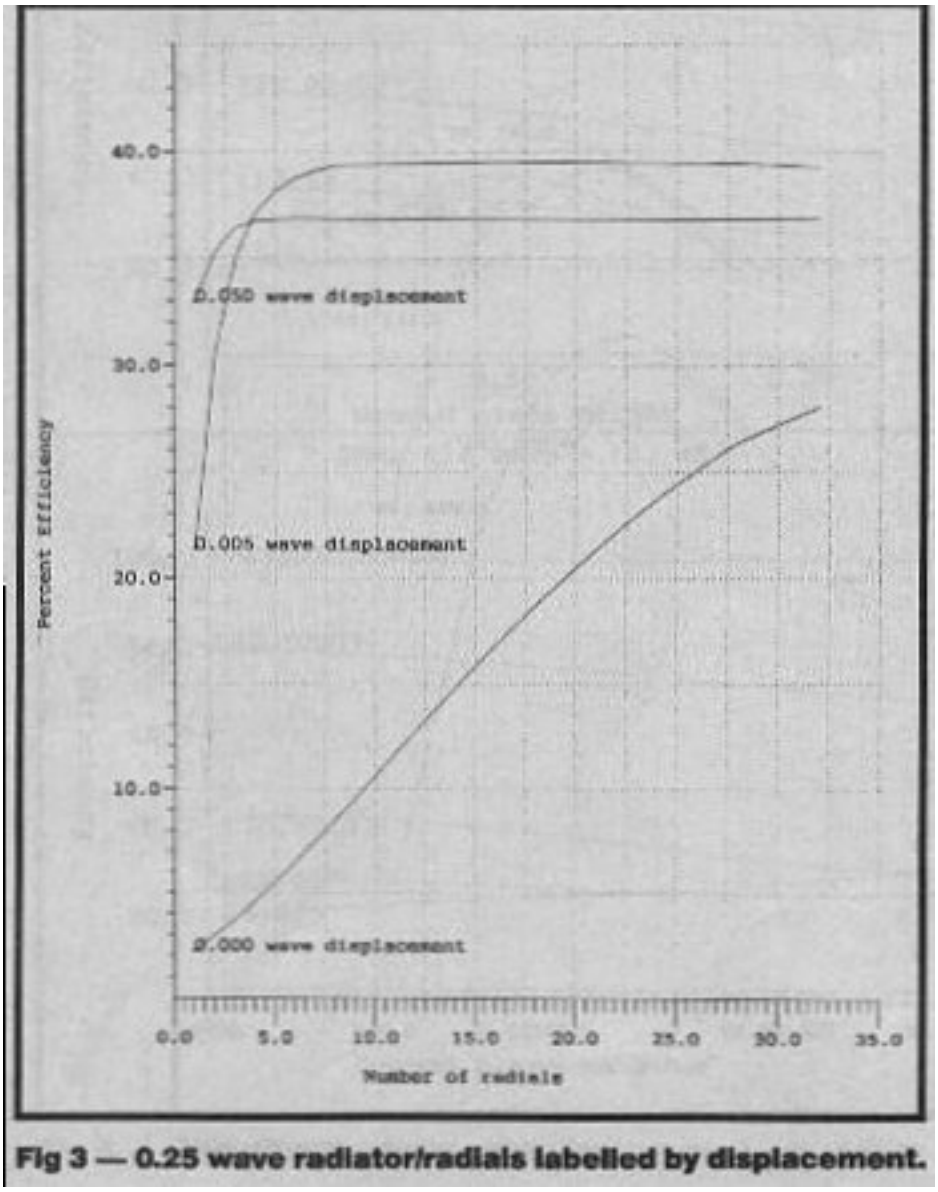


Fig 3 — 0.25 wave radiator/radials labelled by displacement.

Fig 3 - Comparison between tradition and elevated elevated groundplane

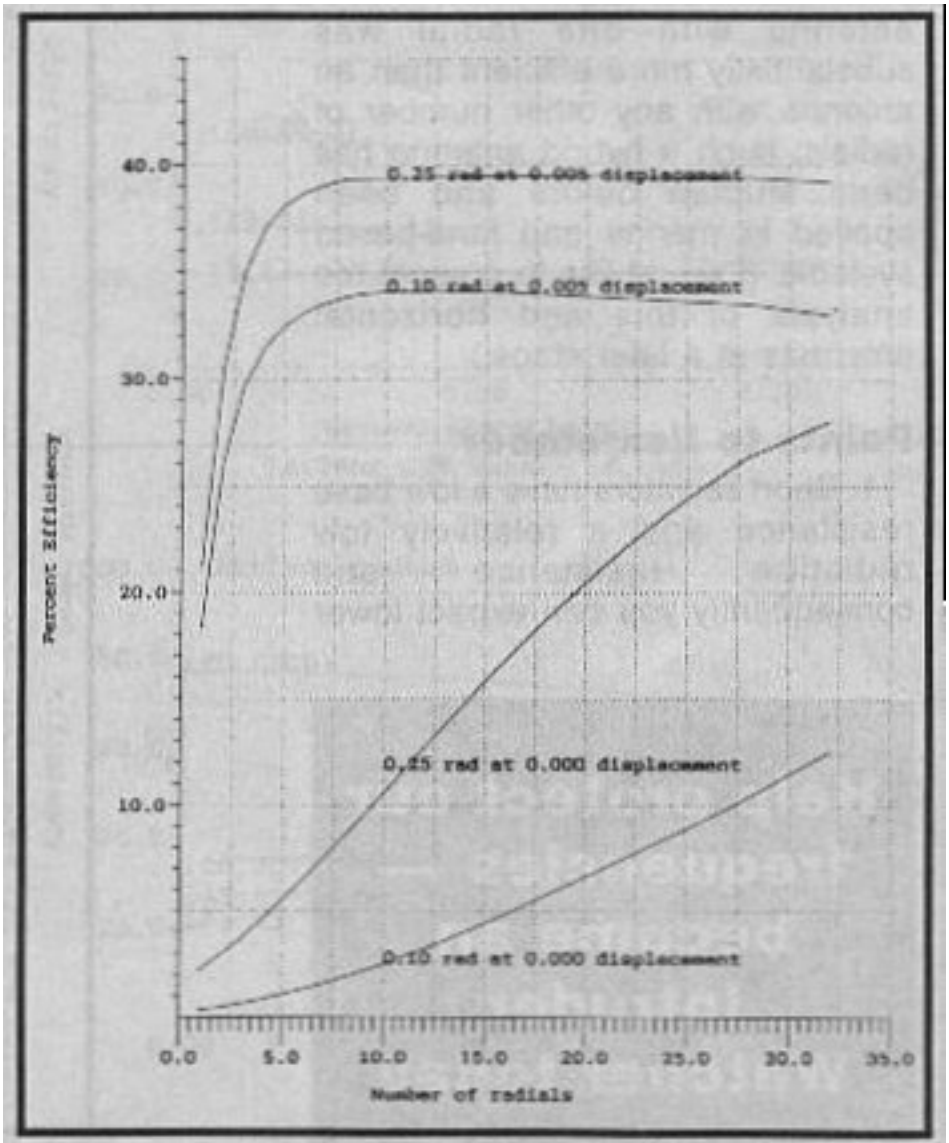


Fig 4 — Radiator/radial length and displacement.

Fig 4 - Efficiency versus number of radials at various heights and lengths

Variable Radiator and Radial Length at a constant Displacement

Figure 5 is the graph for a 1.825Mhz vertical, which was simulated for Eric, VK3AX. This simulation was based on 3 radials mounted at 20' (woolshed height) over typical clay soil (13,5). We were interested in the affect of the antenna height and radial length. The radiator length was labelled on the graph in wavelengths and feet. You will notice that short radiators performed quite well with radial lengths of 0.15 wavelengths and appear worthwhile constructing. When the radiators are lengthened from 0.06 wavelengths to 0.15 wavelengths, the efficiency improves by 5 percent; not much gain results from extending radiators up to the full 0.25 wavelength, in fact, this length was noticeably less efficient than the 0.148 wavelength radiator. In all curves there is an optimal radiator and radial length. Notice how the optimal radial lengths are noticeably less than a quarter wave for the shorter radiators.

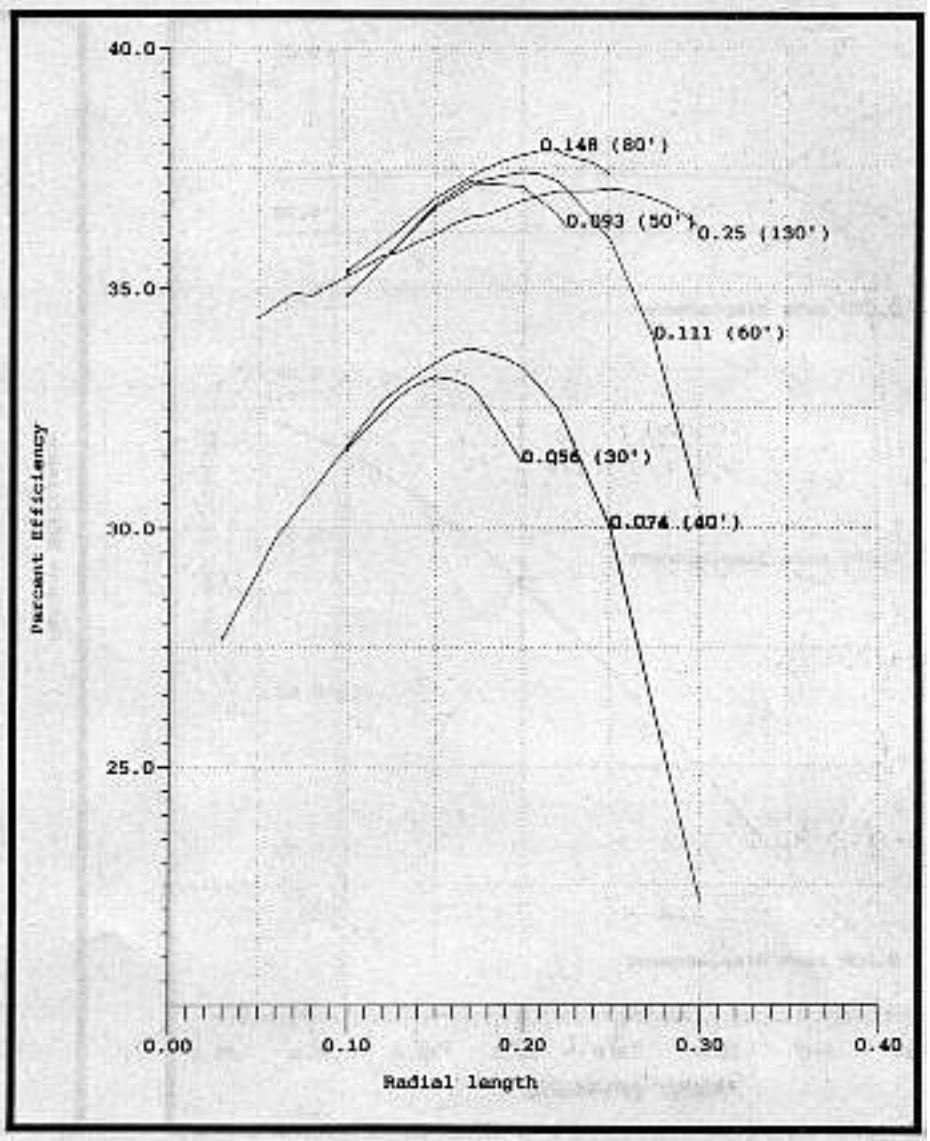


Fig 5 — Fixed displacement at 1.825 MHz 20' (13,5).

Fig 5 - 1.8 MHz antenna with 3 radials at 20'

Efficiency of Elevated Groundplane and Frequency

Figure 6 shows the effects of changing frequency. The analysis has been performed for the 160m, 80m, 40m and 20m bands so you can apply these results to your favourite band. All radiator and radial lengths were 0.25 wavelength, in all cases only three radials were employed.

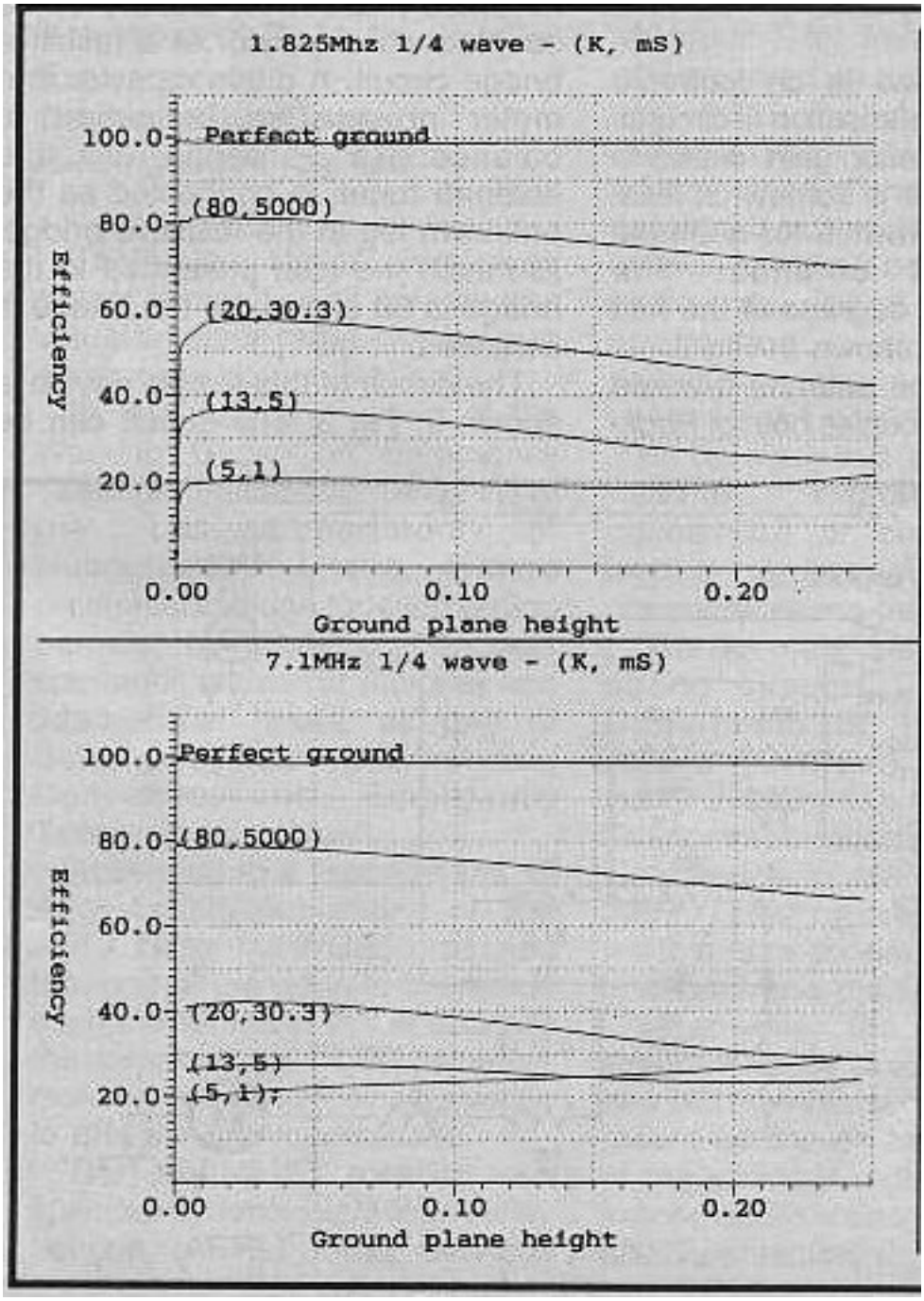


Fig 6A - The effects of changing frequency.

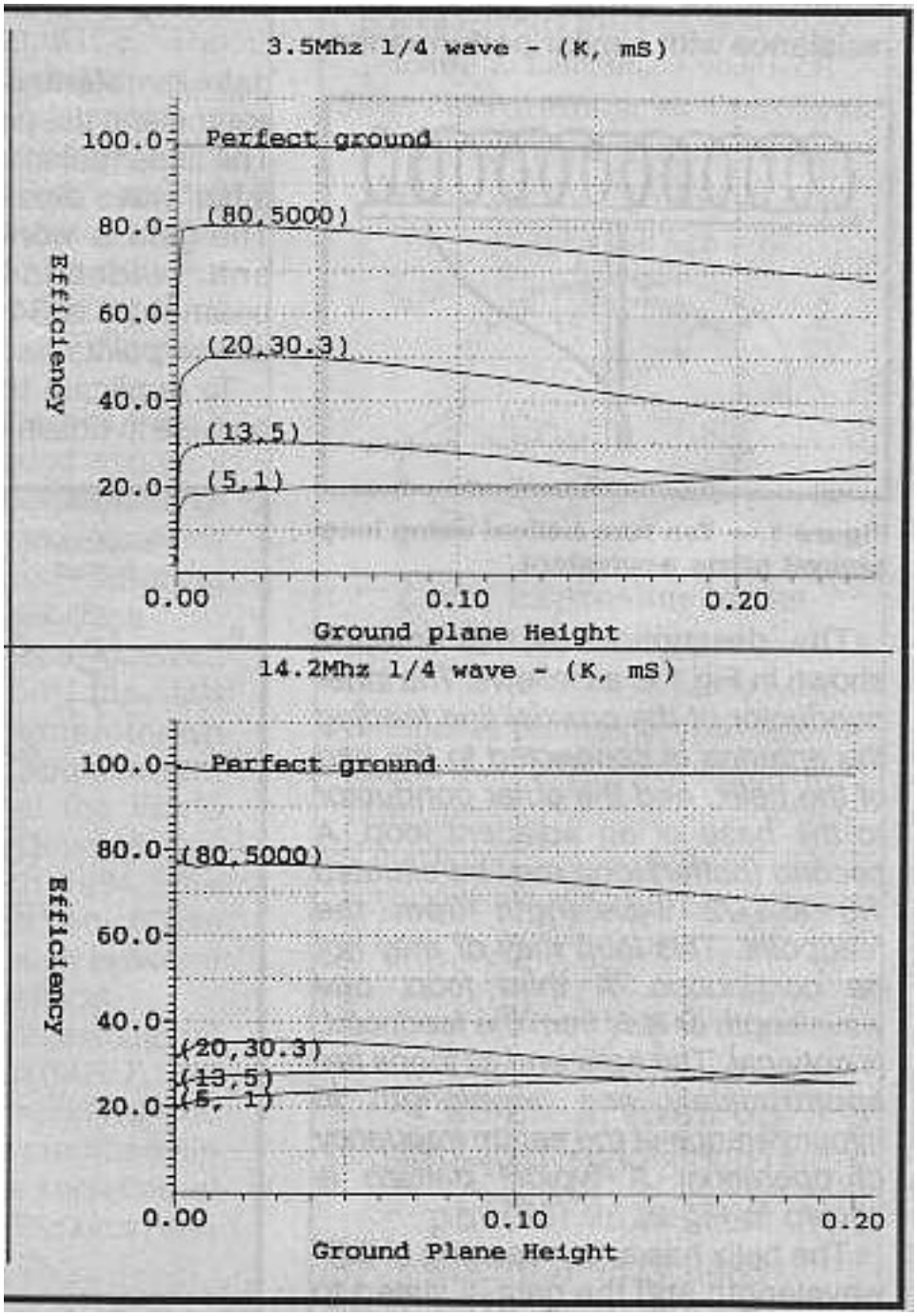


Fig 6B - The effects of changing frequency.

Each curve is labelled by its ground parameters. The curves could be named from top-to-bottom as: Perfect, Sea water, Good, Average, and Poor respectively. Recall that the Average curve, labelled (13,5) is indicative of the dry clay soils.

Notice that the curves peak between 0.002 and 0.05 wavelengths displacement. A more difficult observation, due to the scale of the graphs, is that the curves point steeply towards zero efficiency at zero height (you can obtain a better indication from Figure 2).

Conclusions

From the Figures it is obvious that the simulated antenna efficiency declines dramatically when the groundplane displacement approaches zero.

With three or more radials the radial length is more important to the overall efficiency than the number or radials, with a few exceptions. An interesting side-effect I noticed is that under certain conditions a vertical antenna with one radial was substantially more efficient than an antenna with any other number of radials; such an hybrid antenna has been studied before and been applied in marine and land-based systems. (I would like to present the analysis of this and horizontal antennas at a later stage.)

Points to remember:

1. Short radiators have a low base resistance and a relatively low radiation resistance and consequently you can expect lower efficiency. However, short radiators can still be quite efficient; the shorter radiators require shorter groundplanes and are more optimum at lower groundplane displacements - don't write-off a short radiator! Do take into account the lower feedpoint resistance and consequential difficulties with feeder and element losses, some form of atu at the base is the best. Also note that a short radiator has a very high base impedance, caused by its capacitive reactance, and that the rf voltages at the base are very high!
2. There is an optimal height for an elevated groundplane, that height is not at ground zero, and is typically around 0.05 wavelength. (The statement the higher-the-better is not always true for such systems.) Do not place your ground system on or in the ground unless you have space, time and materials for about 120 radials, unless you are fortunate to have excellent ground, or the desire to hide your ground system. (You may be able to see from Figure 4 that there is a lot to be gained from watering an inadequate ground system - so don't write-off that folk lore!)
3. The efficiency curves are forgiving and somewhat broad; displacements as little as 0.005 wavelengths can be tolerated; some curves actually peak around 0.01 wavelengths.
4. Large numbers of radials are not required for elevated groundplane systems. Three or four radials are sufficient, doubling the number does not double the efficiency.
5. The elevated groundplane system is more efficient at lower frequencies.
6. A lot can be gained by placing your antenna near swamps, lakes and in close proximity to the sea; you should expect substantial improvement, often more than 3db in these cases.

7. Lastly, these results are only as good as the simulation program. However, this program is good at modelling linear antennas close to lossy ground. NEC-2 has been validated numerous times; even so care must be taken because inappropriate use has caused the model to fail.

I have also performed simulations for elevated groundplanes with an underlying secondary ground screen placed on the ground. It is detrimental to connect the elevated groundplane to this form of lower ground system. Note the recommended commercial practice to terminate radial systems with a ground stake - I would highly recommend against such practice; the situation is made even worse if the ground stake is under the feedpoint rather than at the ends of the screen.

I originally thought that this grounding anomaly was an artefact of NEC-2, but I have subsequently read the preliminary publication ***'Recent Advances to NEC: Applications and Validation'*** >by G. J. Burke, 1989, which investigated these effects and others; he used the new improved and classified NEC-4 to model situations for several typical broadcast antennas and antennas below the ground. Burke found, to my surprise, and others, that the recommended grounding practices are detrimental. Burke's findings were based on the relative communication efficiency (RCE) of the antenna measured in terms of field strength at some distance outside the near field, a more appropriate value, rather than the efficiency figure I employed.

Bibliography

The MINI-NEC and NEC programs are based on the Method-of-Moments (MoM); you can find a description of this method in ***" Antennas, Second Edition"***, by John D. Kraus, McGraw Hill 1988.

Another book, which describes the foundation of the technique, is *" **Field Computation by Moment Methods"***, by R.F.Harrington, McGraw Hill 1961. Post publication note: This classic has been reprinted by the IEEE and is available at a discont if you are a member.

The hybrid vertical - horizontal marine dipole, developed by VK3AM, is described in "***Hf Antennas for All Locations***", by L.A. Moxon, G6XN, on page 154.

The *"**ARRL Antenna Compendium***", Volumes One and Two, (now Three, Four and Five) are also interesting reading for antenna construction and modelling.

The book review *"**Computational Electromagnetics: Frequency Domain Method of***

Moments" by Edmund K. Miller, Louis Medgyesi-Mitschang, and Edward H. Newman was extracted and printed in an ACES newsletter. The extract stated that the review contains 528 pages and recommends several books.

One book of recent publication is: "***Generalised Moment Methods in Electromagnetics***", by J.H. Wang, John Wiley and Sons, NY, 1991. I believe this would be for the serious MoM enthusiast.

* Terms

This article was first published in Amateur Radio Volume 63 No 10, October 1995. 'Amateur Radio' is the journal of the Wireless Institute of Australia. . Both the author and the WIA hold copyright. No reproduction is permitted for commercial purposes without express permission of the copyright holders. (mailto). At the time of publication, he author was a member of the VK1 WIA.

This article seems to be the most popular article on my website. The article was updated with post publication notes on 28 Nov 1998, and several hypertext links have been added for your convenience. I modified the article on the 10 Aug 2004 so it uses CSS and because of several requests from readers I have made the graphs scaleable. I apologize for the lack of clarity as the original .jpg files do not scale well. Unfortunately I no longer hold the original data or software from which I generated the graphs.

The document is best viewed with IE 6 and Opera 6. Sorry it may not print properly until I fix the CSS.

This page has been hit **16756** times between 2000-10-01 and 2004-07-30

Usage statistics for all pages.

Quickie Vertical

KQ6RH

(C) 1998, 1999, 2000

Ray Jurgens

(Up-Dated 2/25/2000)

Quickie Vertical (October 1998)

Hear I am sitting in front of the rig listening to everyone working CQ World Wide, and I don't really have time for this, but I hear a few interesting stations on 20 meters and try to call them. Trying to bust the pile up with 100 watts and my hustler dipole is going to be difficult on this band, so I tune up on 15 meters and there is action, and I have a good wire beam, so I tune up and down the band and can't find a clear frequency. Then I hear some one say that 10 meters is open. So I tune up there, and there is a world of DX coming in. I choose a few interesting ones to call, but can't be heard. I should have expected that, as my antenna for this band leaves much to be desired. It is a vertical above my Hustler dipole, but it has no effective ground plane and SWR no better than 3:1 at its best frequency. With no time to assemble a Quad, I start thinking about a better vertical. Suppose I take the parts for the quad and make a good ground plane and put the mobile whip above it. That won't take long. So I quickly assemble a short 4' mast with a center hub and add four 8' long 1/2" diameter fiber glass spreaders and set in on my test stand in the center of the back yard. I dig into my portable antenna box and find an extra mirror mount (Radio Shack variety) and mount it at the top of the mast. Oh, the bolts are too short for the 1" mast, so I am off to the hardware for 2" bolts. I remembered that I had cut eight #16 solid copper wires for a ground mounted vertical about a year ago, so I unroll 4 of them and attache them to the lower bolts on the mirror mount and tape the other ends to the tips of the spreaders. I then tighten the wires by dropping the hub 30" below the mirror mount which pulls the spreaders into the horizontal position and allows the radials to drop at about 17 degrees. Quickly, I mount the Hustler 10-meter whip, connect the coax, and check the SWR. The SWR is great, 1.1:1, but at 29.6 MHz. So I add two more inches to the stinger and recheck. That gave 28.6 MHz for best SWR which is now 1.2:1. Not bad, a total of a half hour has passed, and the same stations are still holding the frequencies. So I start down the list that didn't answer earlier. To my amazement I work each one within a minute. So there I sat wishing I had time to work the contest.

To complete the story, the radials are 101" inches long. The spreaders were at an elevation of 4 feet above the ground. Later, I raised this to a height of 8 feet, and nothing much changed. I also tried resonators for 12, 15, 17, 20, 30, 40 and 75 meters with equally good impedance matches, but I didn't raise them up. The table below gives the measured SWR and bandwidth for each band. A photo of this antenna is shown below the table. As I really didn't design this antenna, we have to call this one just plane luck. By the way, if you want more radials, it is possible to use two hubs with one rotated 45 degrees to the other. Add four more spreaders to get 8 radials. This might reduce the stinger lengths a little, as mine are fully extended on a few of the bands. The 20 meter Ham Stick wasn't long enough to get down into the CW portion of the band, so I added four alligator clips to the tip of the stinger. This worked about the same as the Hustler, So here is an antenna that can be set up in as little as 15 minutes once you have the parts together. If you paint all the parts black, it is nearly invisible at night and not objectionable by day. Note that this antenna works fairly well on 75 meters, but does not work well on 30 and 40 meters. I'll tell you why some time later after I figure it out.

Band (meters)	**SWR**	2.0:1 **SWR Bandwidth**	**Stinger Length (inches)**
10	1.20:1	1.53 MHz	7 1/4"
12	1.05:1	1.00 MHz	11 3/8"
15	1.00:1	1.44 MHz	9"
17	1.00:1	0.52 MHz	20 5/8"
20	1.25:1	293 KHz	15 3/4"
30	5.00:1	--	15 1/4"
40	2.60:1	--	26 1/4"
75	1.00:1	33 KHz	28"

Table 1

SWR, Bandwidth, and Stinger Length for Quickie Vertical, Using Hustler Whips as a Function of Band

Note, the bandwidth is the SWR bandwidth for 2.0:1, and the stinger length is measured from the top of the compression nut. All configurations use the Hustler MO-3 mast part, and all resonators are the low power versions. It is interesting to note the excellent match on 75 meters. Normally, this whip gives a much lower impedance when set on a good ground plane, so this probably means that at least half of the power is going into the ground, even more is lost in the loading coil, but it still works pretty well.

Quickie Vertical Antenna

Fiberglass spreaders stretch out the ground radials. Central hub is PVC. A Radio Shack Mirror Mount supports
a 10 meter Hustler Whip all of which is supported by a Radio Shack tripod mount for testing.

Item Quantity Description

Item	Quantity	Description
1.	1	8' 1" OD fiberglass tube for mast (MGS)
2.	4	8' 1/2" OD fiberglass tube for spreaders (MGS)
3.	1	Hustler MO-3 Mast section and resonator for desired band (HRO) (AES) etc., Ham Sticks also work.
4.	1	Antenna Mirror Mount, (Radio Shack) (HRO) (AES) etc.
5.	4	2" by 1/4" bolts (to replace those in item 4. (Hardware)
6.	1	Versi-Hub (RFJ)
6.	4	1/2" Hose Clamps (Hardware)
7,	4	Terminal Lugs with holes large enough to pass the 1/4" bolts
8.	34'	#16 AWG solid copper wire (Hardware, Electrical)

Table 2
List of Parts for Quick Vertical

Return to Main Menu

Richard Karlquist

"Rick"

Biographical:

Currently employed by Agilent Laboratories, the central research lab of Agilent Technologies in Palo Alto, CA
Agilent is the composed of the former Hewlett-Packard Test and Measurement, Semiconductor, Chemical and Life Sciences divisions.
For previous employment see resume.

Professional Publications:

A New RF Architecture For Cesium FrequencyStandards
A Narrow-Band High Resolution SynthesizerUsing a Direct Digital Synthesizer Followed By Repeated Mixing and Dividing
A 3 to 30 MHz High Resolution SynthesizerConsisting Of a DDS, Divide-and-Mix Modules and a M/N synthesizer
A New Type Of Balanced Bridge ControlledOscillator
The Theory Of Zero Gradient Crystal Ovens
A Low-Profile High-Performance Crystal OscillatorFor Timekeeping Applications

Patents:

4,785,415: Digital Data Buffer and Variable Shift Register
5,148,122: Atomic Beam Frequency Standard Having RF Chain With Higher Frequency Oscillator
5,708,394: Bridge-Stabilized Oscillator Circuit and Method
5,729,181: High Thermal Gain Oven With Reduced Probability Of Temperature Gradient Formation For the Operation Of a Thermally Stable Oscillator
6,651,488: Systems and Methods of Monitoring Thin Film Deposition
6,668,618: Systems and Methods of Monitoring Thin Film Deposition
6,686,777: Phase Detector Having Improved Timing Margins
6,765,519: System And Method For Designing And Using Analog Circuits Operating In The Modulation Domain
6,765,435: Phase locked loop demodulator and demodulation method using feed-forward tracking error compensation

Patents pending: (after 18 months, the USPTO publishes patent applications)

US20030045261: Frequency Translating Devices and Frequency Translating Measurement Systems with DC Bias Added to a Mixer Diode.
US20030053178: Frequency Translating Devices and Frequency Translating Measurement Systems That Utilize Light Activated Resistors.
US20030063626: Method and Apparatus For Synchronizing a Multiple-Stage Multiplexer.
US20030065988: Circuit and Method For Adjusting the Clock Skew in a Communications System.
US20030202543: Aggegate Processing of Information During Network Transmission.
US20030203722: Method of Reducing Power Consumption in a Radio Receiver.
US20030210108: Lumped Element Transmission Line Frequency Multiplexer.
US20040120441: Systems And Methods For Correcting Gain Error Due To Transition Density Variation In Clock Recovery Systems..
US20040119624: System And Method For Divisor Control In A Modulation Domain Divider.
US20040119514: SystemS And Method For Correcting Phase Locked Loop Tracking Error Using Feed-Forward Phase Modulation.
Note: approximately 8 other patents are pending, but not yet published as of March 2004.

Rick's N6RK Amateur Radio Web Site:

Visit www.n6rk.com to see Rick's ham radio web site.

Genealogy of "Karlquist" family

The name "Karlquist" is not very common.
From time to time, I receive inquiries from others with the same name wondering if we are related.
The suffix "quist" means "branch" or "twig" in Swedish.
Unlike "Karlquist", most names ending in "quist" start with the name of a tree
E.g., "Alderquist" (Alder), "Lindquist" (Linden)
I have listed some genealogical info in case you are named Karlquist and are curious.
I know I have some distant cousins in Sweden who go by the name "Carlqvist."

Contact Rick

Send email to Rick
Send snail mail to:
PO Box 2010
Cupertino, CA 95015

10 Meter Vertical Dipole

I was in need of an antenna for my 10 meter beacon, so I decided to build a design I remembered seeing in an issue of QST several years ago. A friend of mine built one at the time and it worked great. It is a simple rigid dipole made from two lengths of 1/2" electrical conduit that are separated by a 5/8" wooden dowel that is inserted into the ends of the conduit and held together with hose clamps. It is necessary to cut two 2" or so long slits in the end of the conduit that the dowel rod is inserted into. This allows the hose clamps to compress the conduit onto the dowel rod to hold it all together. The dowel rod I used is four feet long, so I had about 2 feet in each leg of the dipole for rigid support.

The dipole can be fed directly with coax, or a balun can be used. It can also be fed with twinlead or ladder line. It is important to seal the insulating dowel with epoxy to keep it from rotting. I used the hose clamps to attach the coax to the antenna.

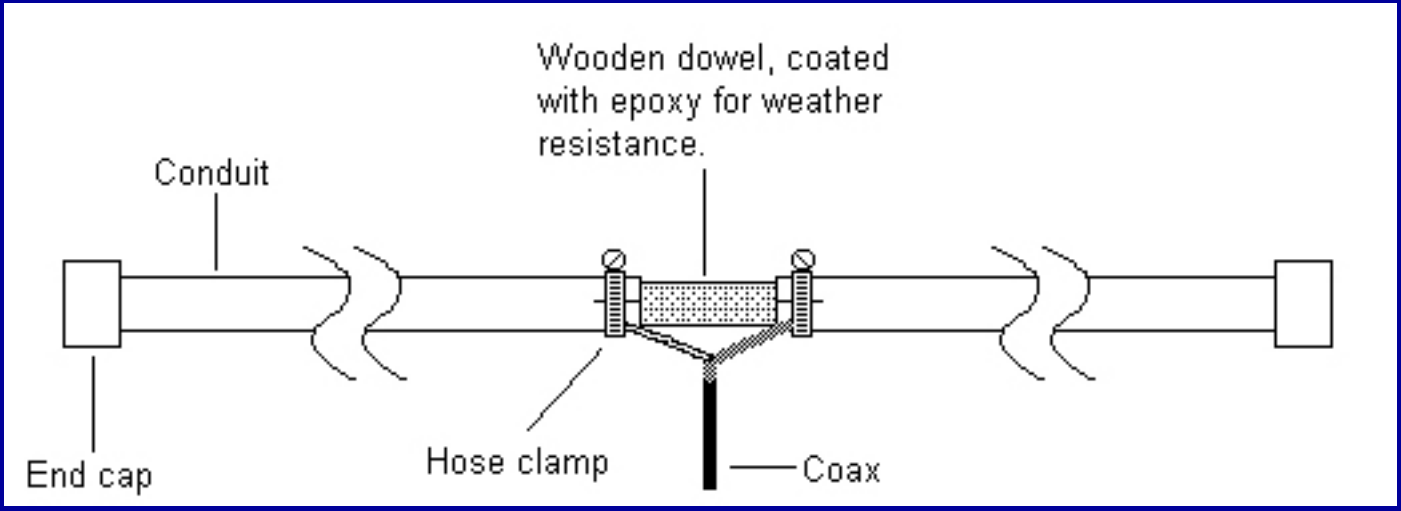
Caps should be placed on each end of the antenna to keep out moisture. I found that the threaded 1/2" PVC caps work well on 1/2" conduit. They can be just placed on the ends but I fixed them on with epoxy. Make sure that this is done after shortening the antenna for best VSWR match.

The length for each leg of the dipole is found using the formula L=234/F, where L=length in inches and F=frequency antenna is to be used on. If the antenna is too short, it can be lengthened by increasing the space between them on the insulator

and lengthening each side of the coax or ladder line feed on the dowel.

This antenna must be mounted so that it is insulated from ground. Mine is mounted on a 4X4 and attached to a 2X4 that is buried two feet into the ground. I used U bolts to secure the antenna to the 2X4. This antenna could also be mounted horizontally off the side of a tower.

This antenna could also be built for the 12 meter band, and possibly even for the 15 meter band. However, most conduit you find in hardware stores is sold in 10-foot lengths, and a 15 meter antenna will need to be around 11 feet long. Some method If lengthening the antenna will be needed. It could be done by attaching another foot of conduit to each end by using connector pieces for conduit.



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No. 26: When Should I Use a Vertical on 10?

L. B. Cebik, W4RNL

Most of the antenna we have discussed in this column have been horizontally polarized. There are some good reasons for this fact. First, 10-meter horizontal antennas are fairly compact, with a half wavelength being about 16-17' long. Second, the shortness of a wavelength on 10 meters (35') generally simplifies the process of supporting a horizontal 10-meter antenna at a good height (at least 1/2 wavelength, with over 1 wavelength preferred for best performance). Third, even 3-element 10-meter Yagis are fairly light-weight, for easy support, even in field or hilltop operations.

Nevertheless, there are some good reasons for using a vertically polarized antenna on 10. Although the gain of such antennas may not usually compete with a well-installed horizontal antenna of the same size, this factor is rarely a problem when the band is open. So let's look at the question of when to use a vertical.

1. **Mobile in Motion**: The standard these days for mobile-in-motion operation is the short, center-loaded, magnetic mount vertical set on the car roof. Although the least efficient of almost any antenna used on 10, these antennas acquit themselves well. Full size 1/4 wavelength whips have gone out of vogue, especially with the increased use of plastics in autos. When auto bodies themselves become universally plastic or fiberglass, we may have to rethink the center-loaded mag-mount vertical for mobile operation.

2. **Lunch-Time Operation**: With small rigs, short antennas, and an open band around noon, 10-meter lunchers are more numerous than we imagine. Since the lunch hour (or half-hour) is all too brief, operators want a system that wastes no time in set-up and take-down. The vertical--again, usually a mag-mount antenna in the parking lot--fills the bill.

3. **Local Convention**: In some towns and cities, most of the locals may use vertical antennas. Sometimes, this represents a lot of mobile work; sometimes it represents former citizen's band operators who have joined the amateur ranks and cut down their old antennas to resonate on the higher frequencies. Since local work is mostly point-to-point, as in VHF operation, cross- polarized antennas result in major losses in signal strength. So if the local group is mostly vertical, then it will pay you to have a vertical at home

(as well as on the car) to join the fun full strength.

Since the path through the ionosphere generally skews signal polarization, distant stations will not suffer from being cross polarized relative to your antenna.

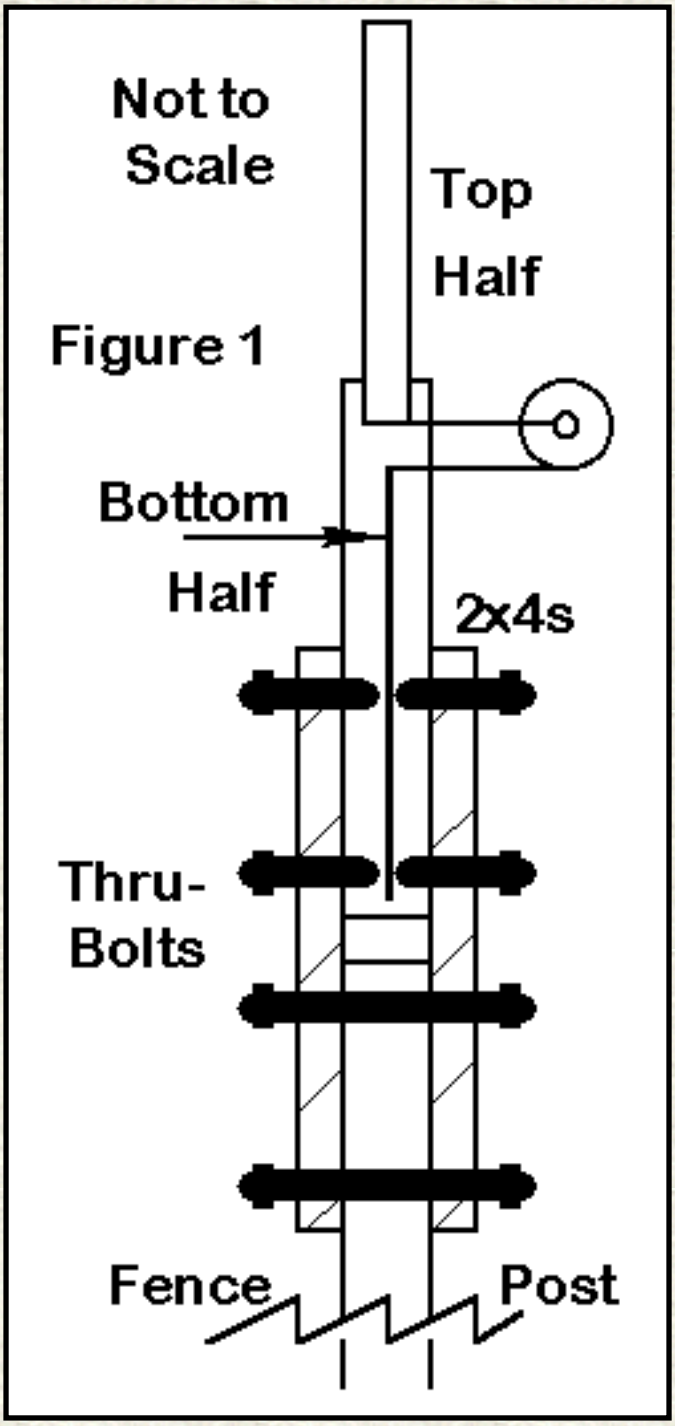
4. **Lack of Space**: Many hams live in homes without large yard. So space for antennas must compete for space with play equipment, patio furniture, and flower gardens. A vertical may be the only antenna type the home owner can erect.

The question here is not whether to use a vertical, but what kind of vertical to use. There are a number of multi-band verticals now on the market that will open many of the ham bands. They come in two major types.

If the roof top is the mounting area of choice, then one of the 1/4 wavelength trap verticals may be best. The heaviest part of the antenna is mounted near the roof top or chimney mounting system for maximum support. The necessary radials, installed according to the antenna makers instructions, can run along the roof top. If the antenna is at the end of the house, radials in the open direction can be run to trees or fence post, well out of reach of children or adults.

Where space is too restricted for an elevated radial system, one of the half-wavelength verticals may be more fitting. Some demand an elevated mounting point and may rest well on top of a fence post, short flag pole, or even a mast attached to a deck post. Other models call for ground mounting and can be placed in the most clear usable place in the yard with buried coax.

In all such installations, safety to children, family members, visitors, and neighbors is a top requirement. These antennas are rarely large enough to cause damage to neighboring property if they fall. Of course, they should be well clear of any utility lines crossing the yard. Finally, they should be isolated so that no one can get an RF burn by touching the antenna while in use. For some models, we achieve this last safety measure by elevating the antenna above reach, even by fence-climbers. Ground-mounted models require some extra thought. Setting up a flower bed and small fence around the antenna can keep most folks away. Sheathing the lower portion of the antenna in large-diameter black plastic down-spout drainage pipe for about 8' up is quite effective in preventing children from touching the antenna and has been found not to adversely affect performance. The protective sheathe can be attractively painted (with non-metallic paint) to call attention away from the antenna. Whatever the safety measures we take, we should also insure that they meet FCC requirements regarding RF exposure to other people.



Even hams with room for a host of horizontally polarized antennas may wish to consider installing one of these multiband verticals. They make good (even if not great) low-band antennas, provide back-up service in case the main beams collapse in high winds or ice, and allow the operator to match the polarization of locals using mobile whips or other vertical antennas. so even if you can afford the highest, the biggest, and the best, one of these simpler antennas makes good sense as part of the antenna farm.

5. **Home Brewers**: Some of us like to build antennas. Some of us have to build antennas to save the cost of commercial versions. Whatever the reason, a vertical dipole for mounting at least 20 to 25 feet up at the center on a non-conducting mast is a good starter project. I suggest a vertical dipole, since it saves a lot of grief over where to run the radials for a quarter- wavelength ground-plane model. The vertical dipole also takes less space than a horizontal dipole and requires no turning for maximum signal.

You can construct a vertical dipole from hardware store materials: aluminum rod or tubing (a little over 8'), PVC, and wood are the main ingredients. **Figure 1** shows in bare outline a vertical dipole I once used to capture Worked All Continents in about an hour at the height of a long-ago sunspot cycle. The 4x4 fencepost was the main support, with underground bracing from bagged concrete. The side rail 10' 2x4s supported a good quality 2x4 mast, with the 4" side running between the rails. Two long galvanized bolts braced the mast. Removing the lower bolt permitted tilt-over operation.

The antenna itself began with an 8' length of aluminum tubing for the top extension. The lower part of the antenna consisted of insulated #12 house wire, purposely cut long. I tuned the antenna to frequency by trimming the lower wire for minimum SWR. Many local hams seemed initially horrified by the idea of a dipole made from unequal diameter elements and trimming only one end. They thought that terrible things would happen to performance, since the antenna was obviously as unbalanced as its builder.

Actually, virtually nothing happens except for a bit of building and adjusting convenience. Half-wavelength antennas lose nothing in performance by being fed slightly (or even radically) off-center. The feedpoint impedance does not begin to change noticeably until the feedpoint is well off center. The only

precaution was for safety: the dipole end is a high-voltage point on the antenna, so it had to be inaccessible to human touch when in operation.

There you have it: some good reasons for using vertical antennas on 10 meters, whether they are commercial multi-band antennas or home brew specials. There are other reasons of a specialized nature that we could add. For example, if you live by the seaside, expect an exceptional increase in performance over the same antenna placed on a rocky hillside in the Smoky Mountains. Verticals have proven to be more than good enough in some island contesting locations. Some operators even prefer the wider beamwidth of a vertically oriented Yagi to one that is horizontal. Whatever the reasons, vertical have and will always have an important place in 10-meter operation, even if we never mention FM and repeaters at all (which I just did).

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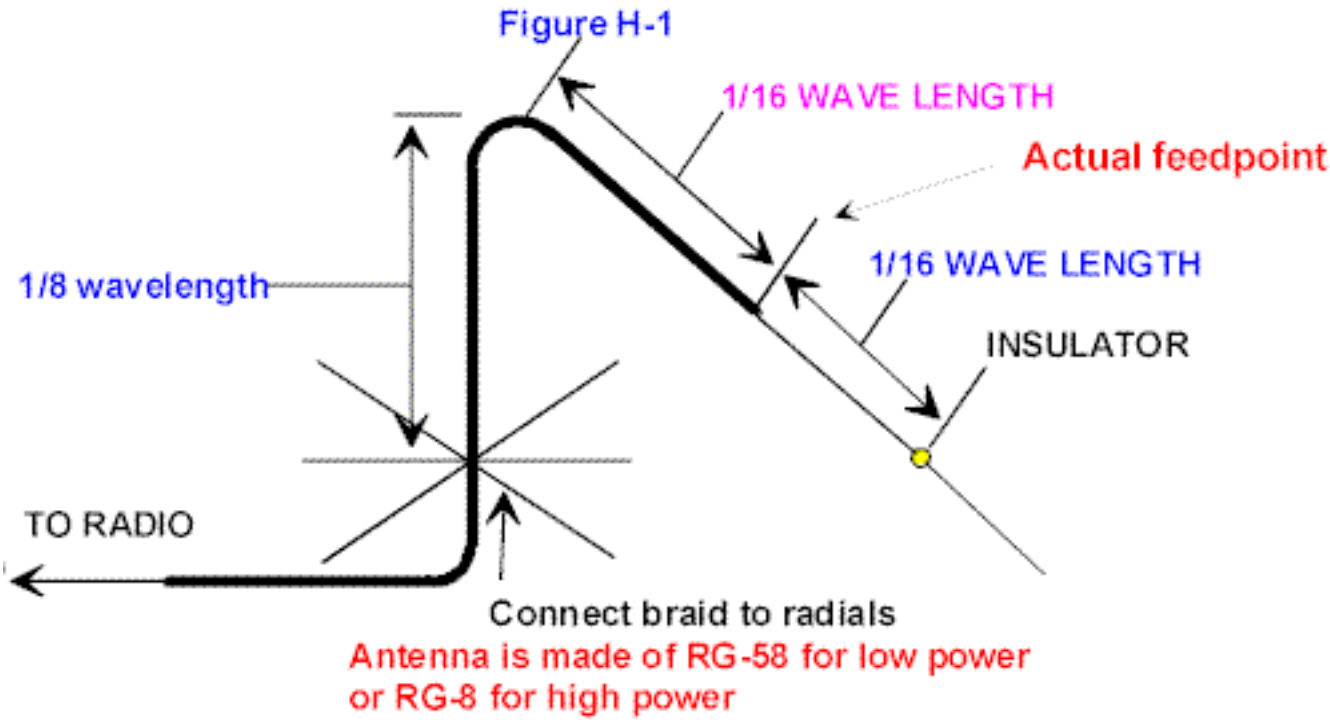
The W5QJR Antenna Revisited

By Richard Morrow, K5CNF



Back in October 1989, the entire paper magazine issue of ***antenneX*** was dedicated to a new antenna concept that engineer/inventor, Ted Hart, W5QJR had come up with in the course of trying to develop a good feed method for his tower on 160 meters. In the process he came up with the antenna that is depicted in Figure H-1. This rather odd looking antenna is a very interesting antenna in that it worked when it was taped to the leg of his existing tower. If it was tuned up there or if it was strung up with an insulating support, it worked just as well. But, an existing tower isn't essential—a tree or similar support will do fine. Here's how Ted's article introduction describes the concept:

"A short wire can be added to an existing tower (or hung from a tree) to form an excellent low band antenna. This antenna ***combines*** the features of a horizontal and a vertical antenna to provide constant signal level to all stations from ***near-in to more than 1,000 miles.*** It is the optimum antenna for rag chewing and net operation on 40, 80 and 160 meters. It is also an excellent DX antenna."

RUNS LOW POWER

So I put one up for 75 meters that was only 32 ft. (9.75 meters) tall. To say it worked is mild. I was only running about 15 watts out and got some reports that were far better than I would have imagined. No one believed I was running so little power. A report of 20 over 9 from Los Angles was the first one I got—then a 30 over from Atlanta, followed by a 25 over from Cuba was enough to convince me that this antenna worked. No only did I get these good reports, I could hear Ted, who lived in Melbourne, Florida, which is about 1,100 miles due east from me across the Gulf of Mexico. He was working into Europe

and making contacts in just about every country east of Moscow. Ted had only a TS-440 on the air, running 100 watts and he was getting good reports. Also, I could hear the stations he was talking to, but could not break into the pileups due to my low power. Moreover, I had a rather large amount of power lines, telephone cables and other utility lines surrounding my antenna. These lines were to the East of my antenna and also very close to the antenna. I am sure that these multiple conductors had a definite effect on my signal in that direction.

SIMPLE CONSTRUCTION

The interesting thing about this is that the antenna was hung from a 35 ft. (10.67 meters) pole held up by a 20 foot (6.10 meter) redwood pole. There was no problem tuning up the antenna, and it worked as long as I had it up. The construction of the antenna is simple and well described in the October, 1989 issue, which is in the Library archives. This antenna is well qualified to be a disguised antenna, since it is so short for the lower frequencies—32 ft.(9.75 meters) for 75 meters, 16 ft (4.87 meters) for 40 meters, and the size goes down from there. An interesting feature is that the folded-over portion of the antenna can be straightened out and extended straight up after going up in frequency past 20 meters. The folded part of the antenna adds a horizontal component to the radiated signal, which enables you to make more reliable contacts in closer than a pure vertical antenna would. This is due to the horizontal and vertical mixture of radiated energy giving higher angle radiation to the signal, yet enough vertical radiation for DX work.

REALLY DISGUISED

This is a very unusual antenna, indeed. It can be put up easily and tuning is not that difficult either. It does work and work well. I have no explanations other than it is another of these antennas that came into being by accident. A very pleasant accident as well, I might add. One thing that is also interesting is that it does not look like an antenna, which adds to the disguisability of the antenna.

Another experimenter who put one of these antennas up was G6RJ, Rob. He had very good results working into the Far East with excellent reports via long path on 18 MHz. For those who are interested, there is several ways to make these antennas directional and if enough interest is expressed, we will put more about this in a future issue.

For now, if you are interested in this antenna, go to the Library under Past Articles and look at the article "The W5QJR Antenna-Revolutionary Concept!" and all of the basics and tuning information will be there and get you on your way to putting up one of these antennas. Many folks have reported finding this to be a very interesting concept and solves some unique problems. **-30-**

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Last modified: April 01, 2001

802.11b Homebrew WiFi Antenna Shootout - 2/14/2

Update 11/2/2003
Added notice at bottom

Update 2/16/2:
I've been Slashdotted!
Some grammatical/spelling errors corrected

Update 2/21/2
How To finshed, linked.

- Introduction
- What the huh?
- The Shootout
- Performance Summary
- The How To

Greg's obsession de' jour

In my efforts to add the words "wireless savvy" to my network admin resume, I've been reading books and web pages on radio propagation, antenna theory and design, and building wireless networks with 802.11 (WiFi). One of the first things that got me excited was the Pringles Can Antenna. Published on the internet and in a fine book by Rob Flickenger, the net admin for O'Reilly, this design for a do-it-yourself, VERY inexpensive antenna made from a recycled junkfood container is as cool as the other side of the pillow. It seems that everyone is building and using these. The various community wireless network groups all talk about them and folks are reporting that they do the job.

A friend of mine built his before me and looking at his finished antenna got me excited to understand the theory of how it works. Reviewing his plan, I came up with different spacing that he Rob did. To see if I could improve upon the design, I built mine with corrected spacing. While waiting for some wireless equipment to come in, I started looking for my next antenna project. Oddly, the more I studied, the less I understood. There seems to be quite a bit of confusion on how the Pringles antenna works and what design category it falls under. The inner lining of a Pringles can looks metallic, but my tests show it not to be. The Pringles Antenna design, and some designs that pre-date it, seem to treat it as though it were metallic. While folks are calling it a Yagi-Uda style antenna, the design of the driven element in the Pringles can antenna looks like a Waveguide style design.

Waveguide antennas don't use the director assembly (the washery bits), and therefore are much simpler to build. An old tin can of the right size, about $5 in parts and 10 minutes of time are all that are needed. The math for computing correct sizing of the components in a waveguideWiFi antenna is simple. Formulas in hand, I started searching my cupboards for tin cans that fit the spec. I found myself staring at the products on the canned food aisle at the grocery store. I even went so far as going grocery shopping with a tape measure. "No no, this spaghetti sauce looks much better. It's about three quarters of a wavelength in diameter, hon!"

What the huh?

On Feb 11th, Rob, posted an article on his newest homebrew WiFi antenna - a tin can waveguide! Rob used a large, 39oz. coffee can and placed a quarter wavelength driven element a quarter wavelength from the back of the can. He reported good results - even better than the Pringles can design used by so many. For the antennas I was building, I was using different measurements based on the antenna design material I had been reading. Now I'm a late entry into this wireless stuff and the experts are going a different way than me. It's time to benchmark.

The Shootout

My plan was to get relative performance measurements for various designs (including mine) of homebrew antennas for 802.11b (WiFi) wireless networks. To do this, I setup a wireless link and changed only the antenna- recording each antennas' performance under identical conditions. I didn't compare them to a commercial directional antenna as my only one has a male connector and I don't have

the right cable to hook it up yet. The contestants were (click on each for design specifications).

- A commercial Lucent "range extender" omni directional
- A buddy's Flickenger-design Pringles can Yagi
- My modified design Pringles can
- A Flickenger design coffee can waveguide
- A coffee can waveguide with corrected radiator placement
- A Hunts 26.5oz. pasta sauce can waveguide that fell in the optimum size range
- A Nalley 40oz. "Big Chunk" beef stew can Waveguide

Performance numbers and methodology

The Performance Summary

The results surprised me! In our test, the Flickenger Pringles can did a little better than my modified Pringles design. Both did no better than the Lucent omnidirectional. Now this is just on raw signal strength, noise rejection due to directivity still makes a directional antenna a better choice for some uses even if there is no gain benefit. The waveguides all soundly trounced the Pringles can designs. I mean they stomped them into the ground on signal strength - as much as 9 dBm better. Every three dB is a doubling in power - that's three doublings (8x increase)!

Of the waveguides, the Nalley's "Big Chunk" took top marks. It was followed by the Hunts Pasta Sauce, my modified coffee can, and the Flickenger coffee can in that order. My three waveguide designs, which utilized the correct theoretical spacing, out performed the Flickenger Yuban coffee can handily. It seems that the design formulas for the waveguide design made a sizeable difference in performance. In the yagis, it didn't matter much. This could be because neither Rob's nor my designs are anywhere near right for optimum performance for a Yagi. I've decided that Yagi design is not for the timid or non-radio-expert.

With these results, I'm convinced that the waveguide design is the way to go for cheap wireless networking. The performance is good, the cost is very low and the skill required is minimal. If you can eat a big can of stew, you can make a high performance antenna.

The How To

Build your own Tin Can Waveguide WiFi Antenna (Cantenna). It's the easiest antenna design I know of.

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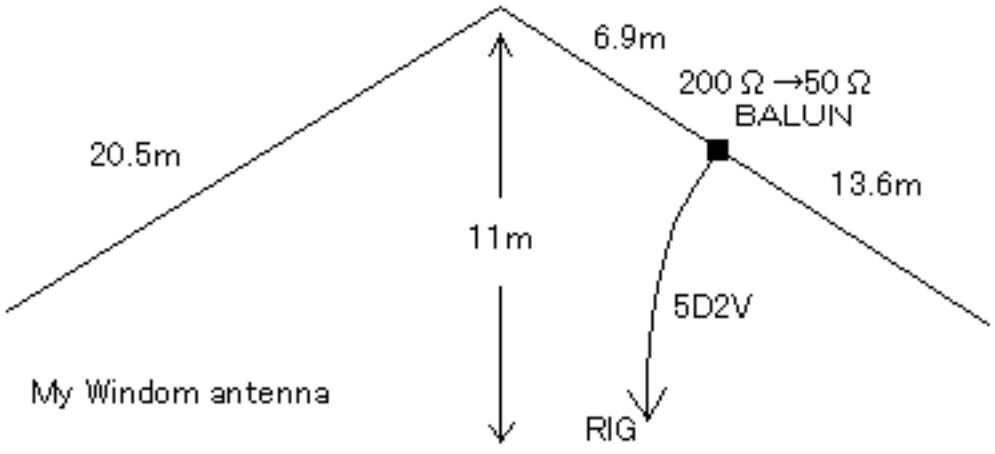
WINDOM ANTENNA

I'm using homebrew Windom-Antenna recently.
In the total length of this antenna is 41m, height is about 11m, and diameter of element is 2mm.
However, it is up only about 4m height from a metallic roof.
I modified this antenna originally used as Inverted-V type of 80m band Dipole a little.
The feeding point of this antenna is located 13.6m from the edge of Element. Yes, it is **Off-Centered**.
BALUN of making which uses two cores, converts impedance into 1/4. The core is the one of about 4cm in the diameter. so there's no problem in the input power 100W at all. Because two cores are used, I think as much as 500W safe in power.
The diameter of the wire rolled in the core is about 1mm.
The VSWR to frequency is shown as follows.

Freq.	VSWR	Freq.	VSWR	Freq.	VSWR	Freq.	VSWR
3.525	1.1	14.050	1.8	18.070	3	28.150	2.5
3.570	1.4	14.100	1.7	18.160	3	28.500	2.0
3.795	1.6	14.150	1.6	21.050	>3	29.000	1.3
7.030	1.4	14.200	1.3	21.449	>3	29.300	1.1
7.060	1.3	14.349	1.2	-	-	29.600	1.2
7.099	1.2	-	-	-	-	29.699	1.3



In 21MHz, VSWR exceeds 3. However it is likely to QSO with the domestic area, if Antenna-Tuner is used. I did not obtain good results though I measured its VSWR with 10MHz and 24MHz besides <showing up>.
I think this easy structural wire antenna is Very FB. Because I can QRV on multi BANDs without switching some antennas.

By the way, I had one question. This antenna might be a name of "Windom" why?...

I thought it might designer's name or name of a place of development ground.

The answer's as following e-mails via radio-amateur's FWD-Net.

> Dear friend, i have a self building Windom; it is a dipole of 80 meter with the connection at 1/3 of the lenght: in this point the impedance is 300 ohm, and with a balun 1:3 i obtain 50 ohm; for the band of the 21 Mhz i put a second dipole only for this band, connected ad the same mode of the other; the lenght of this is 2.3 + 4.6 meters.
> In a test with a vertical groun plane 5 band, the windom result better in 14 7 3.5 Mhz; using a good antenna adapter, it can be used in 1.8 Mhz.
> This antenna is named Windom because was bilded from Laurent Windom (W4DZZ ?) around the 1930.
> It' s not a long time that i use this antenna and i have some thing to understand about it; in any way i'm really satisfate from this antenna.
> I wish to you to have big satisfation using them.
> Best 73 an good DX. Ciao de **Franco. IK1OWB**@IK1JNS.IPIE.ITA.EU

> Hello Tada !
> The "Windom Antenna" was described by Loren G. Windom (QST, Sep 1929, pp 19-22, 84) so this kind of antenna was named after its inventor.
> 73 / **Peter, DJ2ZS**@db0gv.#hes.dl.eu

> The windom is called after its inventor.
> Here I use a windom antenna on 80 - 10 m. the ntenna is 41 meter long.
> 13% off center I feed it with 300-ohm open air-spaced feeder. I use an antenna tuner with the balun coin connected to the antenna tuner. The antenna hangs in zig-zag between some trees. I am very pleased with it, it works quite well locally and on DX. Thought this might be of interest to you..
> Good luck!
> 73 de **Egil. LA8HF**@LA4O.OSL.A.NOR.EU

Thanks to Franco, Peter and Egil. Best regards to you.
And thanks for more infos via the Internet. Those are ...

> I just read an interesting article in Czech ham radio magazine "AMA".
> The article itself was about history of the bobtail curtain antenna, but there was an interesting note: Loren Windom first described this antenna, but he himself said that it was invented by someone else as a modification of Hertz antenna.

I found some people to call the coax fed antenna with a balun "FD4" - but some other people call FD4 the modification mentioned by Franco IK1OWB (but the measures are doubled: 9.4+4.7 meters).
73 **Jindra OK1FOU**

Regarding the Windom, I have a long time experience with such an antenna. I designed mine in 1978. It is different from the "normal" ones because I use a different type of balun and this, from my point of view, is superior to the toroid one because the core does not saturate with high power or reactive loads.
From my point of view, this type of antenna is not rightly called "Windom", it should be called, probably, Long wire off center fed antenna.
The reason is very simple; the Windom antenna was referenced to ground as the feed point was between an off-center side point and ground.
The modern type of "Windom" Antennas are off center fed but not referred to ground. The validity of calling them Windom is due to the idea of "off-center feed".
73 **Gian I7SWX**

The windom antenna should never be configurated as an Interted Vee but should be Flat Top as it was originately.
Matching network such as the Johnson Viking Matchbox is highly recommended
if fed all the way with 300 ohms line.

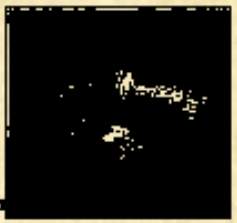
This antenna is suitable for all traditionnal ham bands.
73 from **Cam. VE2SO**

O-kay, Definitely maybe, it should not be called "Windom" that is not single-wire fed or flat-top.
But I will call the antenna for multiband "**Windom**" in high regard.

Well, that is as same as a relation between **H-Hentenna** and Hentenna. You know Hentenna, don't you? Hentenna is a kind of Loop antenna, but H-Hentenna is not loop.
Anyway, **H-Hentenna, that is one of variation of single-wire fed Windom (or 8JK).** It was developed by **JR1FTE** and expanded by **JA7KPI** in 1982. For more infos, **see the LINKs below**.

BACK to SHACK & FIELDS of JA7KPI | **H-Hentenna** for 14MHz | **H-Hentenna** for 50MHz
BACK to FRONT PAGE

Another web page by;

BUX COM ®TM 

"Home of the **RASCAL***"*

My Favorite Multi-Band Antenna

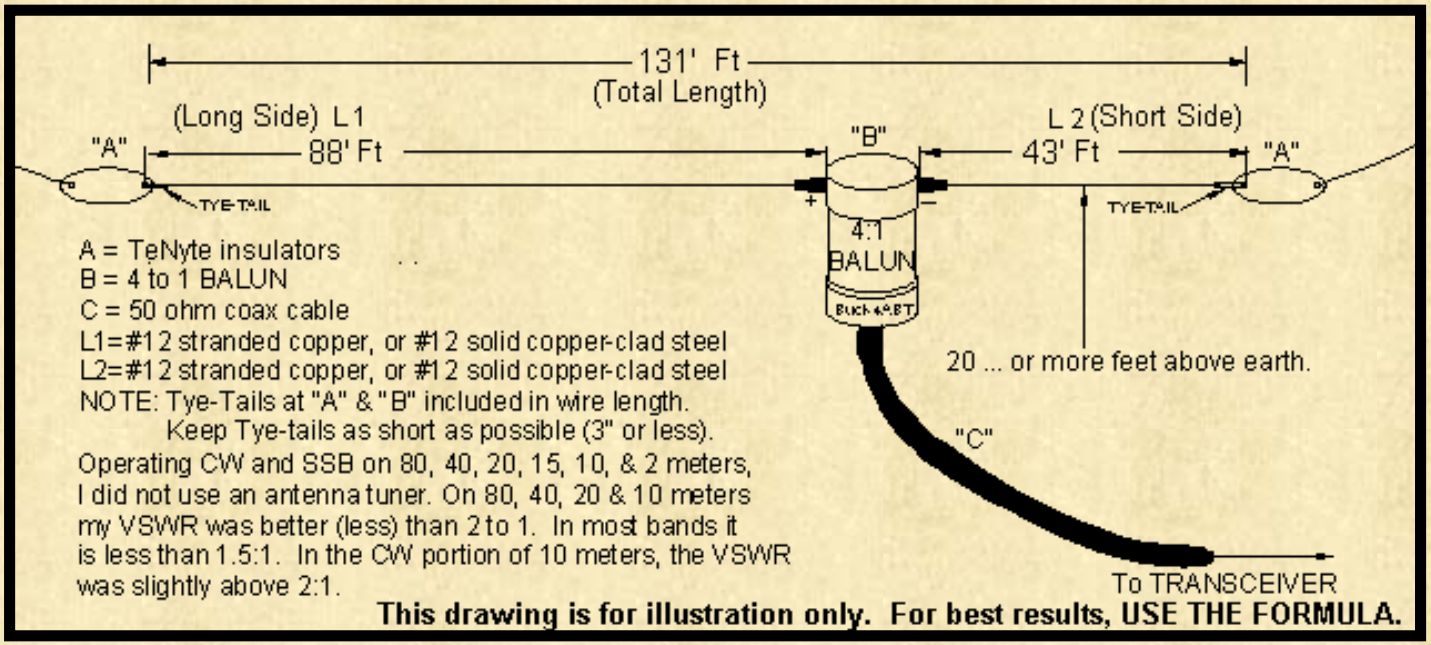
The "WINDOM"

By G. E. "Buck" Rogers Sr; K4ABT

In September of 1949, I was tired of climbing poles and trees to move, remove, add, or change my single-band HF antenna's. In those younger years of my HAM radio hobby, I had used single band dipoles and doublets for almost every HF Amateur band. I had tried long-wires, doublets, dipoles, and Zepps, but again, operation was restricted to single band operation, maybe two bands at most.

I had heard of the "Windom" and read a few articles about the Windom, but most of my thoughts were ... ho-hum.. just another dipole fed a bit off-center. Then one evening at a meeting of the GARC in the old "Sea Scouts" club house near the Coosa River in Gadsden, Alabama; I heard some of my "elmers" Gilbert Watson (SK) W4PAC, Gale Caudle (then W4CFB), Jack Kennamer (SK) W4YPC, Bob Bynum W4USM, Austin (Vic) Vickery, Walter Damkohler (SK) W4EBO, W4CWF, Ed Elkins (SK) W4CDI, Homer Dupree (SK) W4OZK, Jim Runyan, Homer O'Dell, Robert Martin, ... and others discussing the Windom *all-band HF antenna*. It was when Jack (W4YPC), mentioned ***using one (Windom) antenna, on the all HF bands*** *that my ears went directional!*

That last phrase caught my undivided attention. "most all HF bands, ...etc"
What ! *A multi-band HF antenna ?* Surely I had been blessed.



To think that I could put up a Windom, and no longer have to climb the poles and trees to hang another (single band) HF antenna was great news to me. To be able to use it without an antenna tuner was icing-on-the-cake. For a kid without extra funds, an antenna tuner was a luxury that I could *not* afford. Even my transmitter was a single 807 rig I homebrewed on an old *Atwater-Kent* radio chassis, my grand-father had given me.

In those days, a BALUN was unheard of. My Elmer's described, a means of connecting the coax to the off-center fed antenna using a nine (9) turn coil of the coax feed-line at the feed point. This coil of feedline coax formed a "de-coupling" loop. The de-coupling loop provided a crude means of matching the feed coax to the antenna, and at the same time, it would reduce the "re-radiation" (RF currents) along the outside (shield) of the feeder coax.

Today we have toroid cores and BALUN devices that provide a more efficient means of coupling RF energy to the antenna (reducing the VSWR, "standing-waves"), while performing better impedance matching. In the drawing shown above, I've drawn the exact dimensions of the Windom I built in 1949. The only differences in my Windom of 1949 and today are:

1) the material the insulators are made of, and
2) I've substituted a 4 to 1 BALUN for the 9 turn, 8 inch diameter, decoupling loop.

AN UPDATE:

Since writing this article a few decades ago for a major HAM radio magazine, I've received tons of mail (and eMail) asking for more information, especially with regards to a 160 meter version; Here then is "the rest of the story."

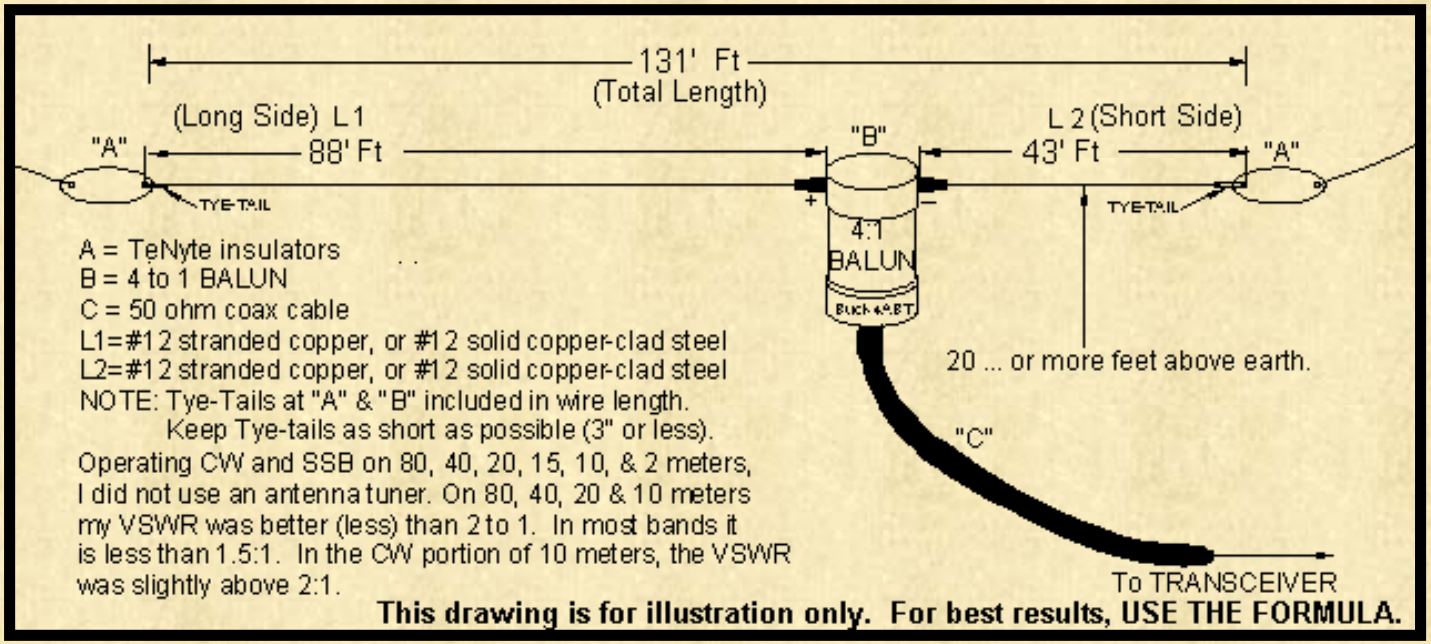
First of all, we'll address the formula, and how to determine the length(s) of each section, using the same old formula that I used in 1949.

Long side.... = 468, divided by the frequency, then multiply by **.**64 (= Feet)

Short side.... = 468, divided by the frequency, then multiply by **.**36 (= Feet)

THUS; for 160 thru 10 Mtrs.....
Long Section; 468/1.8 MHz = 260 x .64 = 166.4 feet.
Short Section; 468/1.8 MHz = 260 x .36 = 93.6 feet.

For 75 thru 10 meters do similar math to arrive at/near the dimensions shown in the drawing below, but for best results, use the formula.

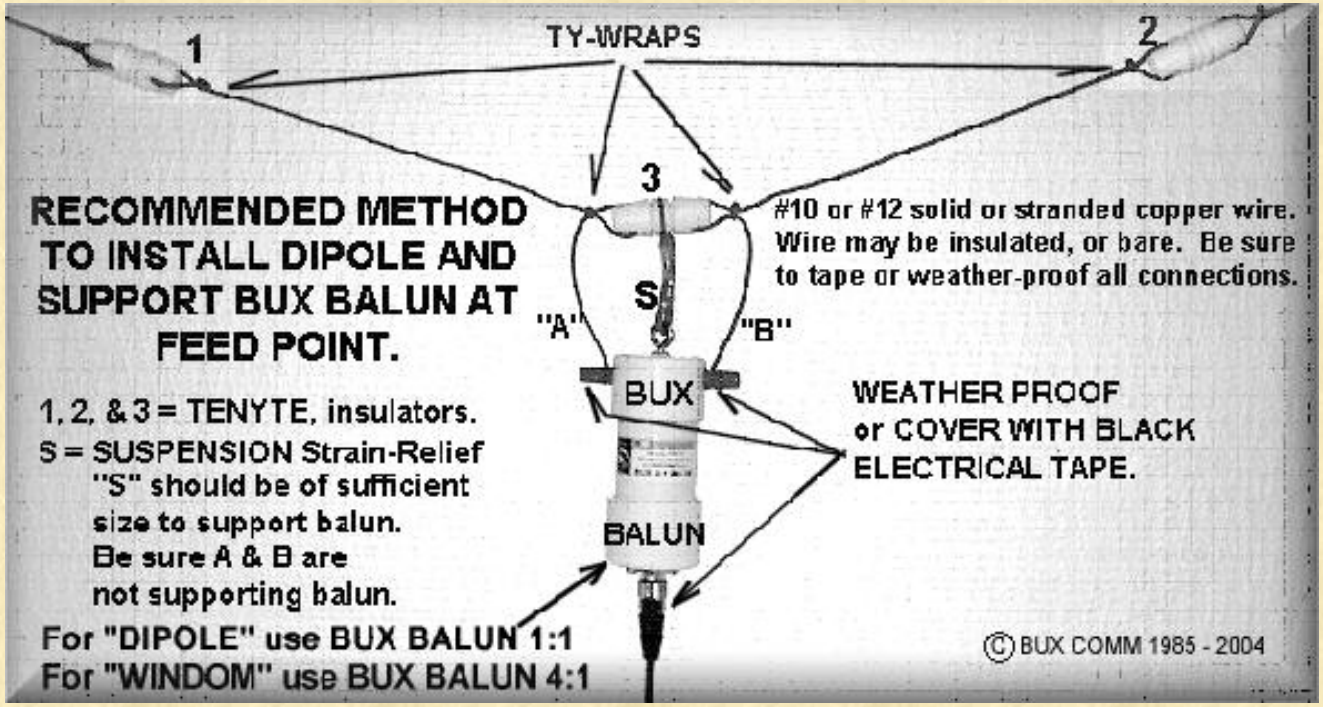


The Windom can be installed as an inverted VEE, or as a sloper, ***but in no case,*** should the angle be less than 90 degrees against itself. *To use an angle that folds against the pattern of the opposite end of the Windom, would change the impedance of the feed-point, change the multi-band features, and most important, destroy the radiation characteristics of the antenna.*

In other words, install the Windom as you would any other dipole, while using a common sense approach. The fact that we are feeding
this Windom using *coax,* and a single balun, gives us the ability to construct it as an inverted VEE or at a "real estate saving (angle)" without destroying the features and performance of the Windom.

This is *not* true with those antennas fed with ladder-lines and those that have several impedance transitions built into the feeders

of the "*basic" Windom..... There's more to come, so read on!*



ANTENNA INSULATORS

Weatherproof, (TENYTE) insulators. Perfect for your DIPOLE or Windom antenna.

For the apartment dweller, you can now hang the 20 meter doublet in the attic. I've QSO'd with stations all over the world with the 33 ft dipole in my garage attic. One insulator at the center, and one each end.

It's great for other HF WINDOM and single-band dipole antennas. Dielectric strength is comparable to the old ceramic insulators, without susceptibility to cracking or breakage under impact or extreme temperature changes.

Package of 3, $2.99 ANTINSL3 Order Now

Package of 10, TeNyte Insulators $8.99, ANTINSLX10 Order Now

Now it's time for me to P O the wire and cable vendors, and the proponents of Windom's with too many feed-line transitions. Twin Lead, Ladder-line, Window Line..... etc.. is an over-kill.. The trade-off is not...; I repeat; NOT... worth the pain and upkeep to replace it every year or two.

And trust me... It is NOT a problem of the ladder-line insulation quality or properties that I'm speaking of. I speak from experience; Wind will destroy ladder/twin-lead line. Even the robust type will succumb to the wind element in time.

IF; you decide to use ladder line, make one turn (twist) to each 20 to 24 inches of line to decrease the wind resistance presented by the "flat" line..... even if it is "window-slotted" type ladder-line. By adding the one twist per 20 inches, it may last longer than three years.

COAX does NOT present a high degree of wind resistance, and it'll last much longer.

Having *been there, done that*, and heard similar horror stories from others; another question arises regarding parallel line currents that come with the use of so-called balanced (twin-lead) feed-lines. To add injury to the ladder-line proposition, the balanced line may also assume the properties of a *single wire feeder (yuk)*.

Some purveyors of the Windom that use ladder-line transitions, must use two impedance matchers (or BALUNS) with their (knock-off) Windom. One is to transition the coax outside the HAM shack *(heaven forbid, we use ladder-line inside the shack... RF "feedback" in everything)*, to the ladder line, then another at the Windom (antenna) feed-point to choke off parallel current from the ladder-line.

Since the feed point of this antenna has been found to be near 220 to 260 ohms, *the use of a 4:1 BALUN to join our coax to the antenna,* emerges as a compromising solution.

Let's not lose sight of the most important advantage of using this antenna; and that is: It provides us with a powerful, multi-band antenna, and a minimum of feedline components.

Now for the "perfectionists" among us, if you want to smooth out the hills and valleys on some of the bands, add your antenna tuner to the system, and the Windom becomes very flat across the bands. When I say "flat," we are talking about reducing the VSWR to a minimum, to produce a good forward power, transfer of RF energy to the Windom.

We're having fun already... with the Real WINDOM antenna

The "Windom Antenna" was described by Loren G. Windom in QST magazine, September 1929. Pages 19 through 22. It is named after its inventor/designer.

Loren Windom, W8GZ, was first to reveal the antenna to the radio amateur community by describing the antenna in the September 1929 issue of QST. It was by Windom's name that the antenna became known. The Windom antenna is an off-center fed dipole with an unbalanced coax feedline.

In 1937, the Windom was first described as a compromise multiband antenna. The antenna can be employed on 80, 40, 20 and 10m with considerable, though acceptable levels of VSWR. What became perhaps the most popular multiband Windom design of all, was the German-made Fritzel FD4 antenna, described by the late Dr. Fritz Spillner1, DJ2KY, in 1971. It had the same dimensions as the multiband Windom antenna, but fitted with a 200Ω (4:1) balun in its feedpoint and fed with coax.

Today, many radio amateurs are using multiband Windom antennas with more than satisfactory results. It would not be without reason that Windom antennas are being employed during IARU HF World Championships! and most of all, by "high-stake-contests." ***Perhaps many young hams ignore the multiband Windom antenna because of its sheer simplicity and may be thinking it is too good to be true.*** *The complexity of feeding other dipoles and doublets, the losses in dipoles with traps and the esoteric marketing of some other antennas seem to appeal to them more.*

CLICK HERE: and read more about the evolution of the WINDOM, to ZEPP, to VHF J-POLE.

BUX BALUNs should be installed at the antenna feed point, or where the coax or feed-line attaches to the above ground antenna. BUX BALUNs are used to connect balanced antennas to unbalanced transmission lines, such as coax cable. Their primary purpose is to prevent antenna (RF) currents from flowing down the outside of the cable. Another function of the BUX BALUN41 is to match the impedance of an unbalanced coax to the balanced feed point of a balanced input antenna(s). BUX BALUNS may also be used as “line isolators” anywhere along the cable to prevent the destructive influence of induced RF currents (VSWR).

BUX 1:1 BALUNs are current BALUNs. They consist of several large, number 73, ferrite type 44 cores. ***BUX BALUN11*** operate from 3.5 to 72 MHz and allow use of RF power to the rated capacity of the BALUN.

BUX 4:1 BALUNs are voltage BALUNs. They consist of a large, number 41, ferrite dough-nut bobbin. ***BUX BALUN41*** operate from 3.5 to 55 MHz and allow use of RF power to the rated capacity of the BALUN.

During this month, when you purchase either of our BUX BALUNs, we will include 3 (Tenyte™), insulators FREE. A $2.99 value. *Be sure too mention this offer in*

the "Comments" space on the order form check-out page!

At BUX COMM, *We don't cut corners!

The components used in the manufacture of our BALUNs are from top quality components, beginning with the Silver Plate SO239 connectors and center insulator is made of teflon™(E.I Dupont). The wire we use to wind the ferrite donut is heavy-duty, silver flashed wire, with teflon™ insulation that will handle RF voltages above 5000 volts, and temperatures above 2000 degrees. The binding posts are heavy-duty, tempered brass, with end holes and side-thru holes to accommodate either type loop-thru connection. A double-shoulder brass capture nut is used to add a secure bite and improve antenna wire electrical connections. With our BUX UN UN (ONION), the coax is Belden™ and the PL259 connectors are Amphenol™.

4:1 Balun, BUXBALUN 41 $19.95 Order Now

o 50 ohm, SO-239 unbalanced input
o Balanced output
o 1.6 to 50 MHz
o Toroid (Voltage) design
o Heavy Duty, Lightweight construction
o Sealed against moisture

1:1 Balun, BUXBALUN 11 $19.95 Order Now

o 50 ohm, SO-239 unbalanced input
o Balanced output
o 1.6 to 50 MHz
o Toroid (Current) design
o Heavy Duty, Lightweight construction
o Sealed against moisture

BUX UN UN *De-Coupling transformer, similar to above,* but has *SO-239 (female) input* connector and *output connector is 1 ft Mini 8 cable with PL-259 (male).*

BUX ONION $19.95 Order Now

4:1, *1.5kw* Balun, BUX BALUN 41HD $27.95 Add To Cart

Toroid design, wound with teflon covered, silver plated wire.* Heavy-Duty, construction.

1:1, *1.5kw* Balun, BUX BALUN 11HD $27.95 Add To Cart

Toroid design, wound with teflon covered, silver plated wire.* Heavy-Duty, construction.

BUX ONION Decoupling Transformer

BUX UN UN De-Coupling transformer, has SO-239 (female)input connector and output connector is 1 ft Mini 8 cable with PL-259 (male).

BUX ONION is and UNbalanced to UNbalance (UN-UN) *decoupling transformer* designed to be used specifically with **the DBLSPCL antenna.**

High RF currents traveling along the coax feed-line shield can cause high VSWR. This decoupling transformer prevents RF currents from traveling down the outer shield of the coax.

The input connector is an SO239 (female) and the output connector is a PL259 (male), which mates the connector of the "DBLSPCL" RV/Apartment dweller antenna shown above.

BUX ONION $19.95 Order Now

WHY USE A 4:1 BALUN

Krusty Olde Kurt is now going to repeat himself. Why? Because the same question keeps coming up over and over. And he wants everyone to get it right.

"I'm feeding my dipole with 600-ohm line. At the station end I need a balun to convert to 50-ohm coax. I need a 12:1 balun, right?"
Wrong! *A **4:1 balun** would be better.*

Why is that? If your dipole is up, let's say, 35 feet then on 80 meters it will probably have a resistance at resonance of about 40 ohms. The actual resistance depends on the height above ground in wavelengths.

If the dipole is 40 Ohms then what do you see at the transmitter end of your 600 ohm line? If the line is a half-wave long (120 ft on 80 Meters) you'll see 40 ohms. Remember, a half-wave line repeats what it sees at the other end. But if it is a quarter-wave long you'll see 8500 Ohms! At other line lengths you'll see impedances somewhere between these two extremes.

So you are not going to see 600 ohms at the end of your 600-ohm line. That only happens if you have a 600-ohm antenna hooked onto it. With such a variation in impedance at the trans-mitter end of the line there is no one balun transformer that will match it. Most of the time the impedance will be above the 50 Ohms of your coax so a high impedance balun would be desirable. Unfortunately high imped-ance baluns don't work well when not matched.

Experience has shown that ***4:1 baluns work best*** in this service. They are more rugged and will take bad mismatches especially if they are wound on an iron powder core. So stop searching for that 12:1 balun. **Use a 4:l BALUN and your system will work just fine.**

You can read Kurt N. Sterba "AERIALS" column in **World Radio Magazine.**

BUX MOBALUN

When you hear those strong HF mobile signals, here's one reason they stand out from the rest.
By installing the BUX MoBalun near the input to your antenna, you deliver more RF energy to the antenna. At the same time, the BUX MoBalun prevents RF from traveling back along the shield (high SWR) of your coax. High power rating, Low-permeability toroid, with Internal composition fiber-glass to prevent vibration during mobile operation. For input and output connectors, we use only the best Silver plated, Teflon insulated, SO239 connectors. 700 watts PEP. *Our **MoBALUN** is also ideal for marine antenna installations*.

$19.95 MOBALUN Order Now

ANTENNA INSULATORS

Weatherproof, (TENYTE) insulators. Perfect for your DIPOLE or Windom antenna.

For the apartment dweller, you can now hang the 20 meter doublet in the attic. I've QSO'd with stations all over the world with the 33 ft dipole in my garage attic. One insulator at the center, and one each end.

It's great for other HF WINDOM and single-band dipole antennas. Dielectric strength is comparable to the old ceramic insulators, without susceptibility to cracking or breakage under impact or extreme temperature changes.

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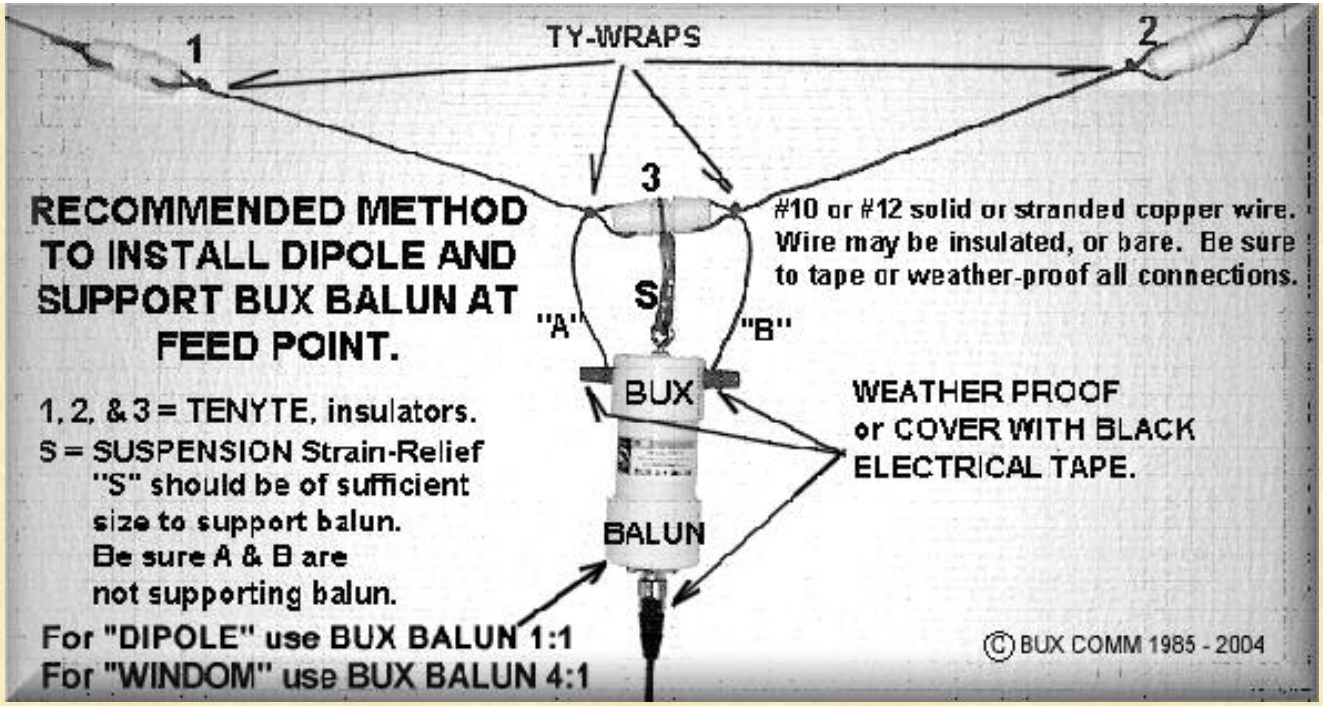
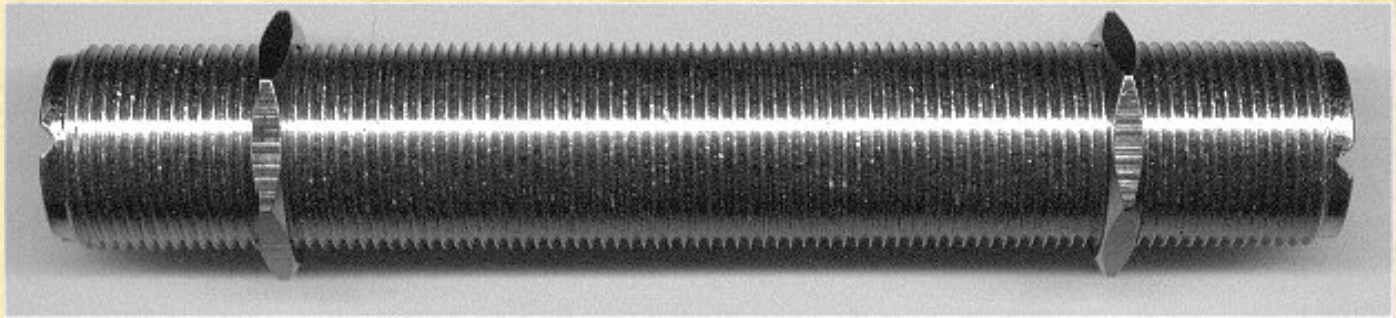
Package of 10, Insulators $8.99, ANTINSLX10 Order Now



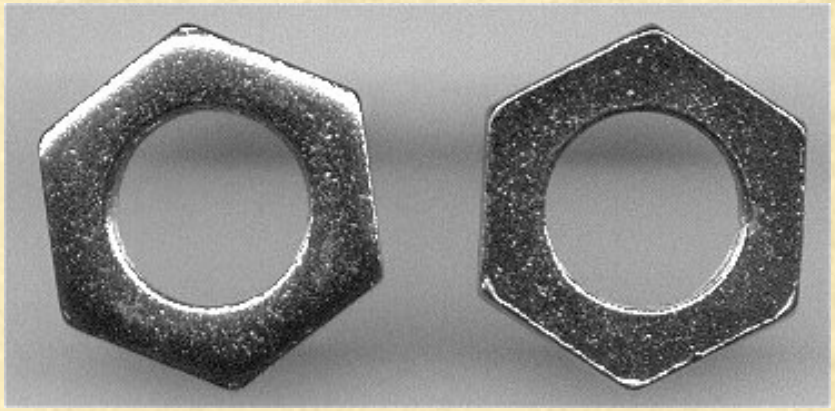
UHF DOUBLE FEMALE BULKHEAD (feed-thru) CONNECTORS



For bulkhead and through-the-wall UHF connector feed-thru connections, with keeper nuts.:

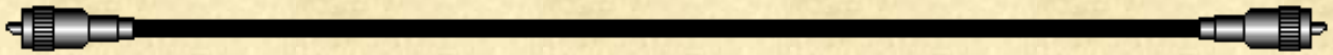
Order 7518-2 (Two inches)	**$ 2.95 ea.**	Order Now	**10 for $ 24.50**	Order Now
Order 7518-4 (Four inches)	**$ 3.95 ea.**	Order Now	**10 for $ 34.50**	Order Now

Order 7518-6 (Six inches)	**$ 4.95 ea.**	Order Now	**10 for $ 44.50**	Order Now
Order 7518-8 (Eight inches)	**$ 8.95 ea.**	Order Now	**10 for $ 82.50**	Order Now
Order 7518-10 (Ten inches)	**$ 10.95ea.**	Order Now	**10 for $ 99.50**	Order Now
Order 7518-12 (Twelve inches)	**$12.95ea.**	Order Now	**10 for $ 114.50**	Order Now



Heavy Duty (1") Nuts for the above bulkhead connectors.

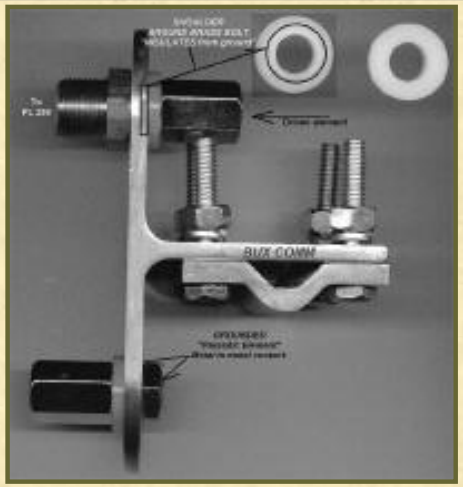
2/.99 cents HDM1-NUT Order Now **10 for $ 3.99** Order Now



50 Ohm impedance, 3 ft, 6 ft, 9 ft, 12 ft, & 18 feet.

These jumpers are made from high quality, low-loss, coax cable with PL-259 connectors installed at each end.

Three ft. (3')	**ORDER No. 8X3-PLPL $3.95**	Order Now	
Six ft. (6')	**ORDER No. 8X6-PLPL $4.95**	Order Now	
Nine ft. (9')	**ORDER No. 8X9-PLPL $5.95**	Order Now	
Twelve ft. (12')	**ORDER No. 8X12-PLPL $7.95**	Order Now	
Eighteen ft. (18')	**ORDER No. 8X18-PLPL $8.95**	Order Now	



Reinforced, Double Bracket used with our "DBLSPCL" antennas shown elsewhere on this page.

Mount two mobile whips in a horizontal plane to form a compact apartment dweller Dipole, or RV antenna. Comes with two sets of mounting hardware (bolts), those shown here and another set of longer bolts for larger masts. Double Antenna Bracket as shown at left.

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After a few years, the hardened, nylon shoulder washers become weather-brittle and break. Keep a few spares in the Ham Shack:

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By combining a pair (two) of the single band HF Mobile (loaded) tuneable-tip antennas, and (VHF) whips at left, you can have an ideal, single element, horizontal dipole (beam). This type antenna opens the world of HF and VHF communications for the apartment dweller, RV HAM, field-day operating, and a variety of other uses and applications. Shown mounted on 1 & 1/4" mast. Antenna(s) Not Included.

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By combining a pair (two) of the single band HF Mobile (loaded) tuneable-tip antennas above, you can have an ideal, single element, horizontal dipole (beam). This type antenna opens the world of HF communications for the *apartment dweller, RV HAM, field-day operating,* and a variety of other uses and applications. The packages described below include two antenna, the Double Antenna-Bracket (DBLBRKT), and mounting hardware

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6 Meters → Add to Your SHOPPING CART

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K3MT
and daughter KF4LGR
present . . .

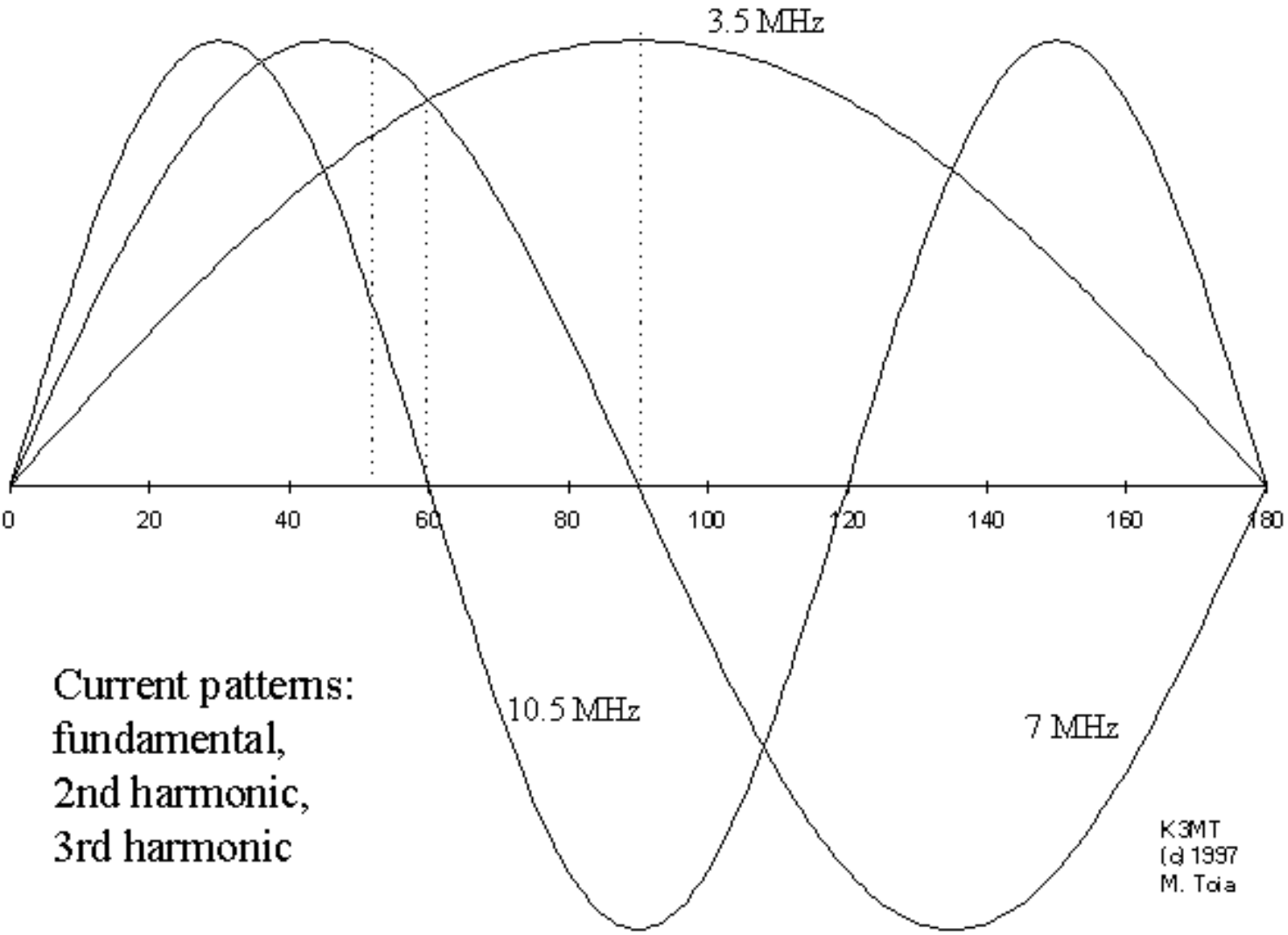
A six-band, HF Windom antenna

April, 1997
Corrected September, 2000

This Windom antenna was marketed in the late 70's and early 80's as *Smithe's Windom.*
It was designed to cover 80, 40, 20 15, and 10 meters. By serendipity, it also covers the 17
and 2 meter bands.

(Authors' note - we now market a revised version of this antenna that covers all HF bands except
30 meters, and covers 30 meters with use of an antenna tuner. e-mail us for details.)

Now, how was a Windom antenna developed? It began with a center-fed, half-wave dipole.
This antenna also works fairly well on all *odd* harmonics, because the center of the
antenna has a current maximum, just as a half-wave antenna has. But on *even*
harmonics, the center of the antenna has a current *minimum.* It is a high-impedance,
center-fed Zepp antenna on even harmonics. This figure shows the current standing wave on a
3.5 MHz half-wave dipole, and the currents on the second and third harmonics (7 and 10.5 MHz.)



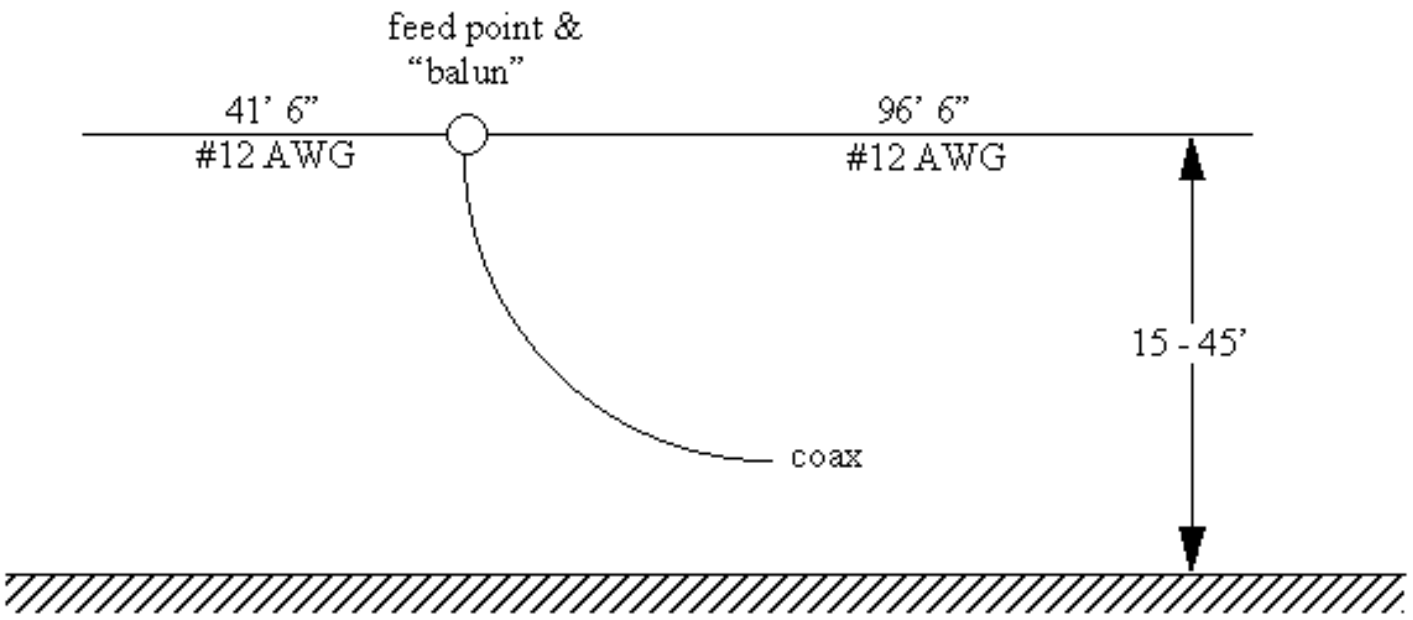
When fed at the center - 90 degrees from one side - a good match to coax occurs on 3.5 MHz.
But the match at 7 MHz is bad: the current is a minimum, so the impedance is very high.
So try feeding it 60 degrees from the left end. Since the current at 3.5 MHz is lower than at
the center (and the voltage is higher) the feed impedance is higher - over 100 ohms. But the
antenna is still resonant, so the *reactance* is low! What you have done is to
increase its feed resistance.

Look now at the action on 7 MHz. The feed point is no longer at a current

**minimum. Therefore,
the second harmonic feed impedance is quite a bit lower than it had been earlier, and is in the
range of a few hundred ohms. Since the antenna is resonant here, too, it has low reactance.**

**But now the feed impedance at 10.5 MHz is poor, because the 3rd harmonic current standing wave
is now a minimum. So try feeding it at about 52 degrees from the left end. Here the match at
3.5, 7, and 10.5 MHz is fairly good. The impedance at all three is now somewhere around 200
to 400 ohms.**

Now you can play these games all day, and if you build this antenna,

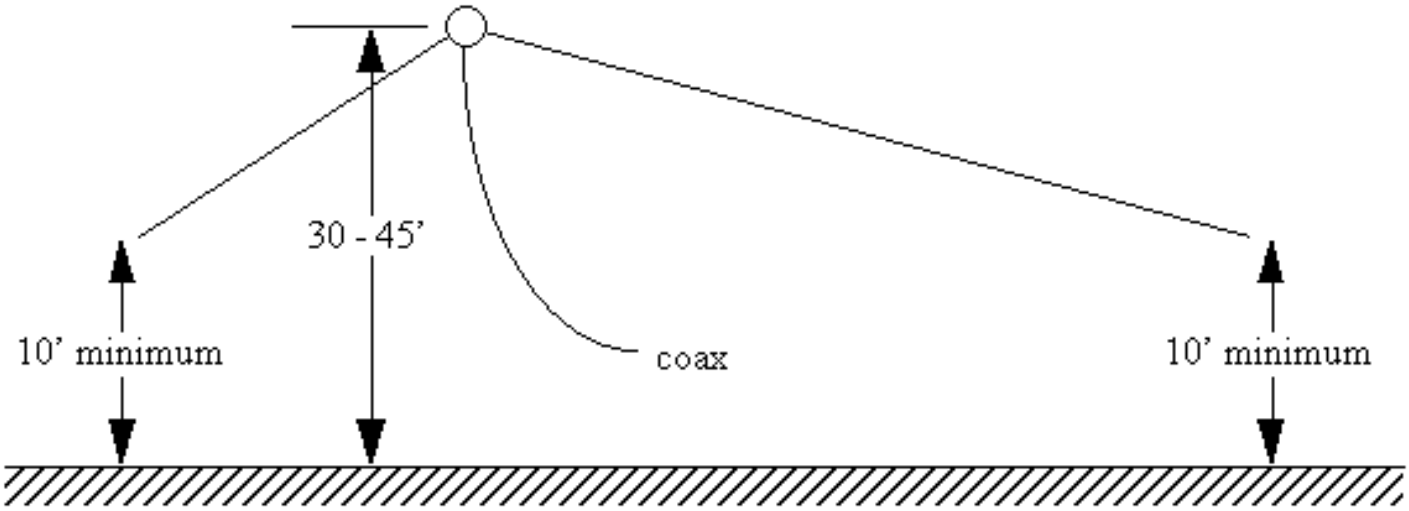


Off-center fed Windom
mounted horizontally

K3MT
(c) 1997
M. Tola

you will find it works well on 80, 40, 20, 17, 15, 12, and 10 meters - plus 2 meters as well,
provided **you pay attention to the balun! To boot, the balun matches 50 ohm coax**
without an antenna tuner. I admit, that this is a compromise design, and a tuner helps on the
low end of 80 meters a bit, and on the high end of ten. But without a tuner, and with a fussy
rig - my Drake TR-7 - a lot of DX has been worked on all bands, from 80 through 10 meters.

I put my windom up a bit differently, as shown here:



Mounting the Windom
as an inverted vee

K3MT
(c) 1997
M. Tola

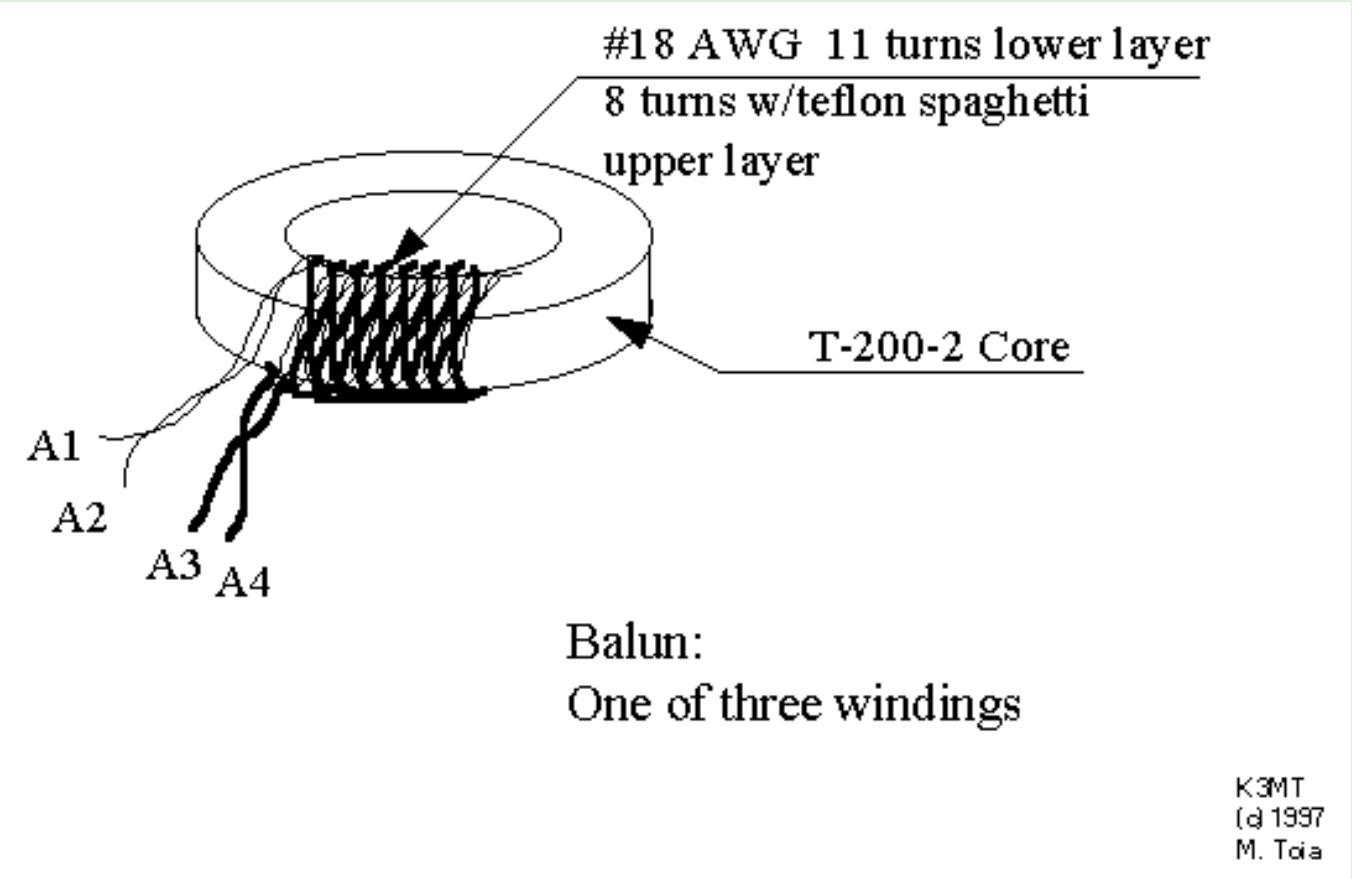
Balun Details

What about the balun? The original unit sold with the Smithe Windoms is a Guanella-type (as opposed to a Ruthroff-Sevick design) parallel transmission line balun. Since the design impedance was measured to be between 300 and 600 ohms, a 9:1 down-converting balun with three 150 ohm lines was designed and built.

To build one, obtain an Amidon T-200-2 core, tape it with two layers of black poly electrical tape, and obtain some #18 AWG magnet wire with a bit of #17 AWG teflon spaghetti. Twist the magnet wire to make three twisted pairs - about one twist per inch. Wind 11 turns of one pair on the core, and slip the teflon spaghetti over *each* lead of the remainder (untwist it a bit to do this.) Then wind 8 more turns back overtop the 11 turn winding. Do this with the other two twisted pair lines as well. Space them on the core so no two lines overlap.

(Apologies and thanks to readers who noted that the original web article called the core a "T-250-2." The correct core is a T-200-2.)

This image shows a single winding on the core - make two more windings like it.

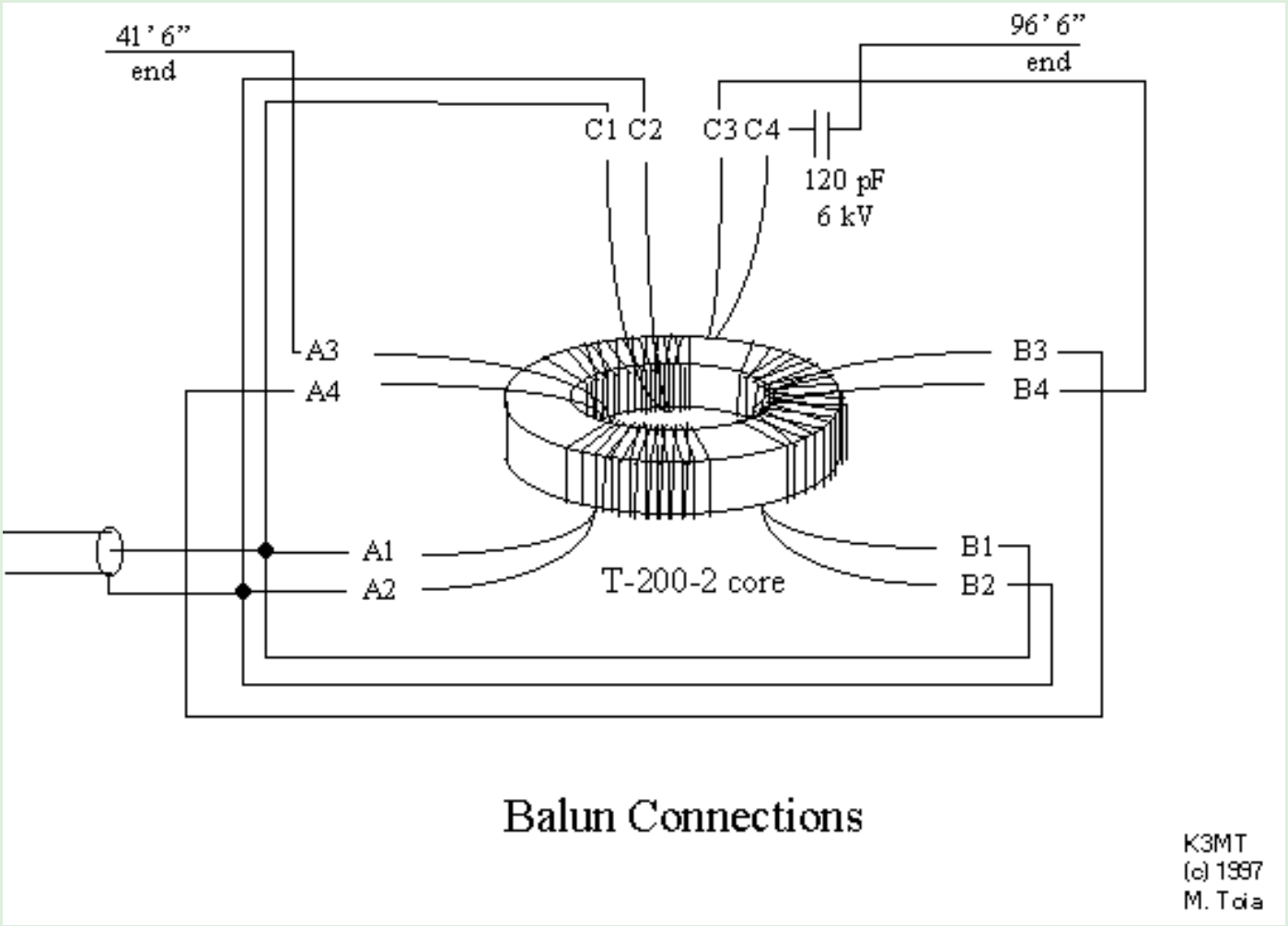


Get an ohmmeter to check continuity. Label the lines A, B, and C, and their ends 1 & 2 where
the uninsulated wire starts onto the core, and 3 & 4 where the wire (insulated with the spaghetti)
leaves the winding. Pay attention to the wiring detail that follows, and use your ohmmeter to
check your work. Label the wires so there is continuity from:

- **A1 to A3**
- **A2 to A4**
- **B1 to B3**
- **B2 to B4**
- **C1 to C3**
- **C2 to C4**

Refer to the next image to guide the balun connections, and wire the balun as follows:

- **Connect A1, B1, and C1 together. These will connect to the center conductor of the coax.**
- **Connect A2, B2, and C2 together. These must connect to the coax braid.**
- **Connect A3 to the *short* end of the windom. *This is important!***
- **Connect A4 to B3, and B4 to C3.**
- **Connect C4 to a 110 pF, 6 kV capacitor.**
- **Connect the other end of this capacitor to the *long* end of the windom.**



NOTE: if you think the balun is too complicated to build, e-mail us. We can

provide one.
Ask for pricing. In fact, we will be glad to provide an entire windom, according to this
design, if you wish.

You are now ready to install and enjoy your windom. If you have the same luck that K3MT and
daughter KF4LGR have, it will have been worth all the trouble!

For more unusual antennas, visit our web page.

This antenna is featured in K3MT's Book about HF antennas

73
K3MT
KF4LGR

X-Beams

KQ6RH

(C) 1998, 1999, 2000

Ray Jurgens

(Up-Dated 2/25/2000)

X-Beams

The X-Beam has been described in QST and the ARRL Antenna Compendium # 1 in enough detail that it should not be difficult to assemble this antenna and obtain the specifications indicated here, though the specifications claimed by the author may be a little over-stated (see reference 1). The claim is that the antenna has 6.0 dBd and 18 dB F/B ratio. The figures below indicate the reality of the situation based on the E&M modeling program. I present both the free space patterns and the ground mounted pattern for 20' height.

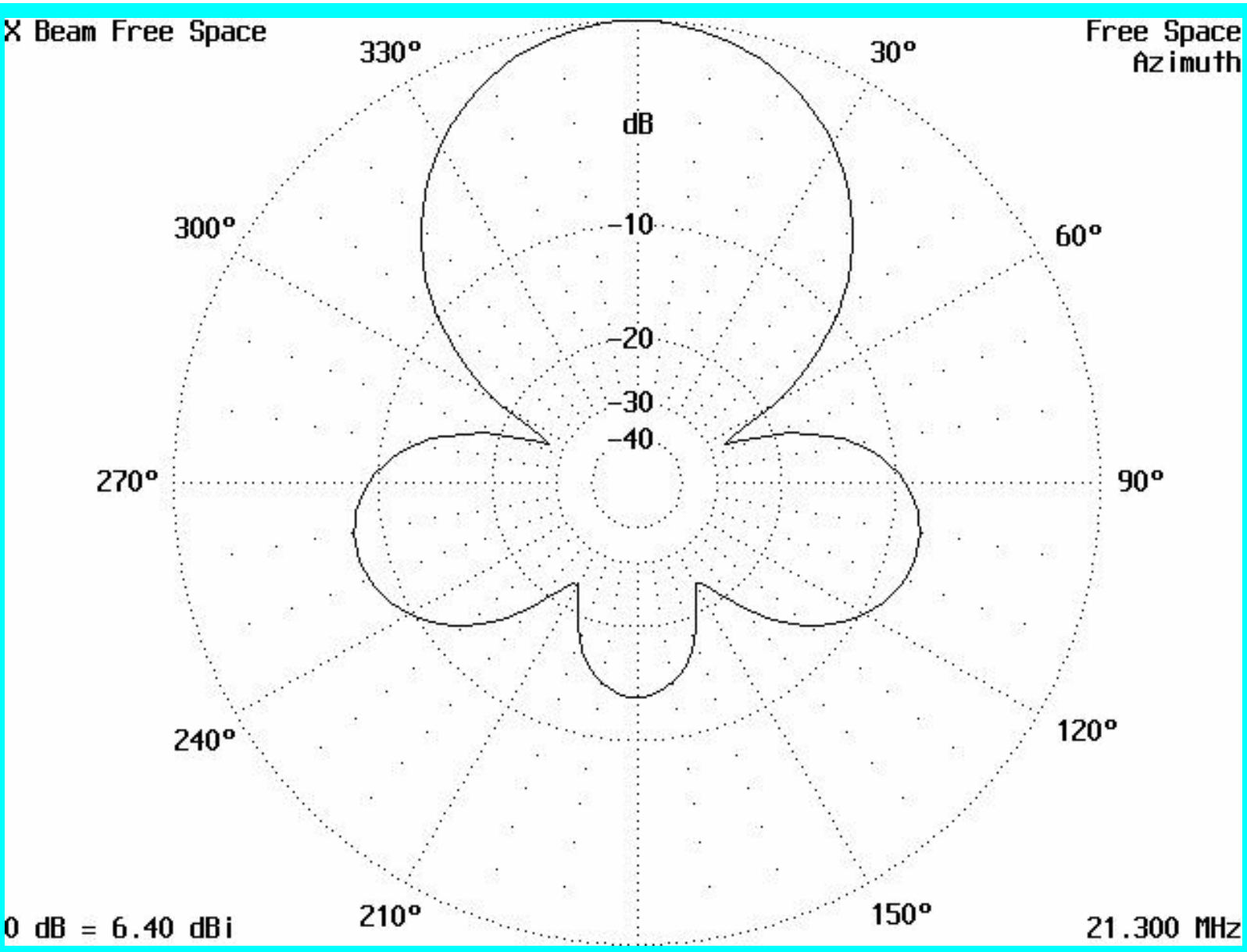


Figure 1

Azimuth Pattern for X-Beam in Free Space

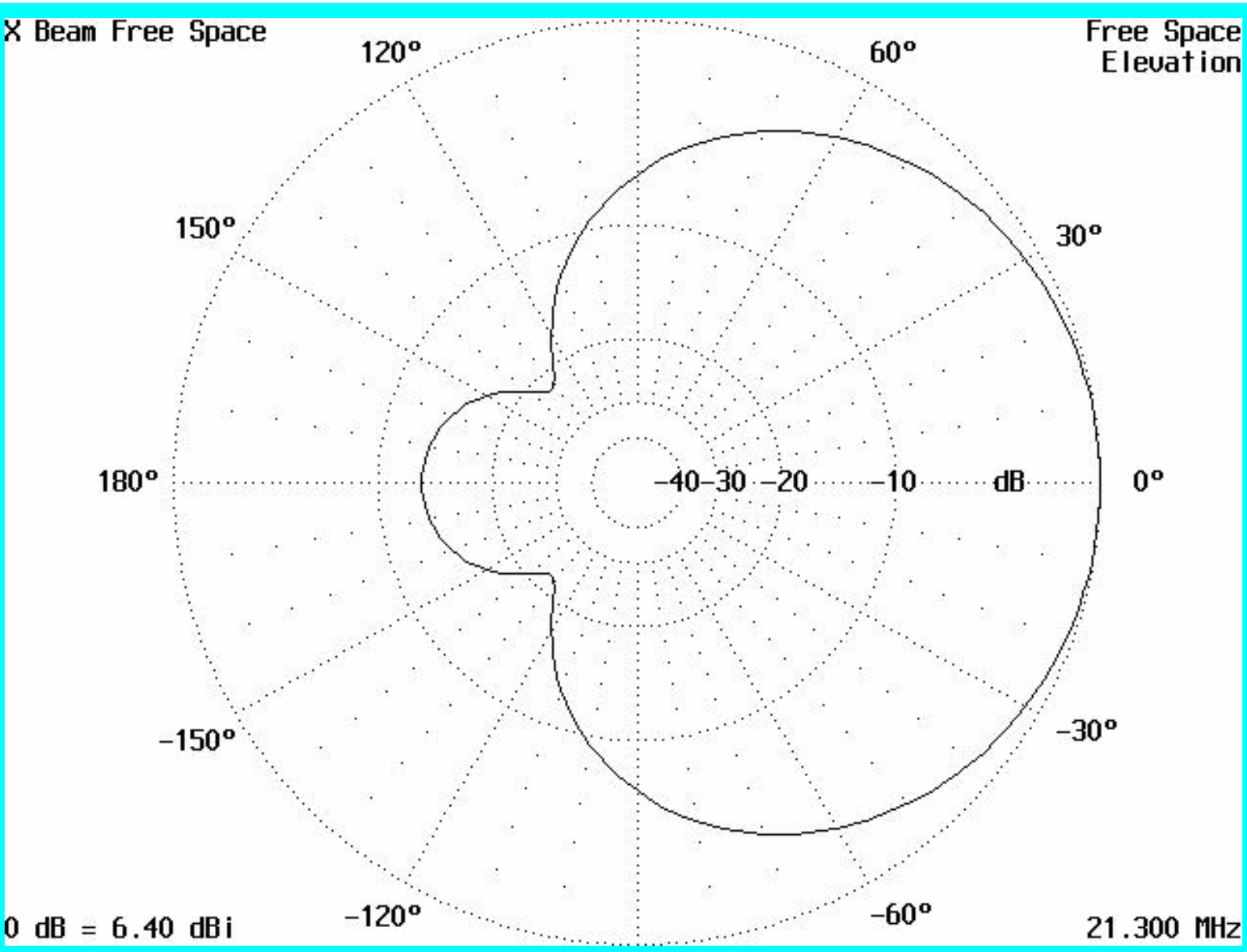


Figure 2
Elevation Pattern for X-Beam in Free Space

Note that the total gain is 6.4 dBi, not 6.0 dBd, and although the F/B ratio is about 13 dB, the side lobes are much worse (about 8 dB). The performance of the antenna looks a bit better when placed about 20 feet above the ground as indicated by the following two figures.

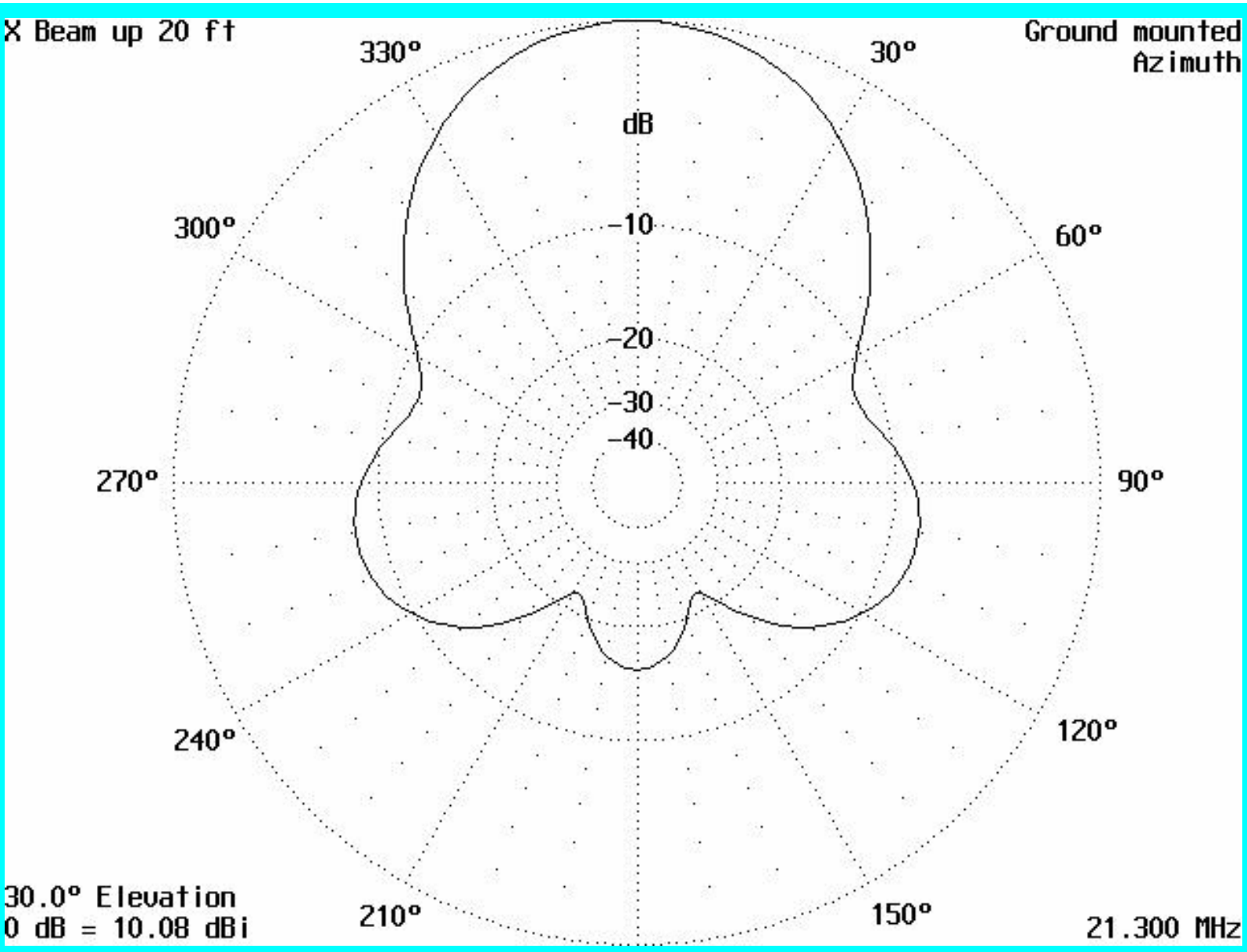


Figure 3

Azimuth Pattern of X-Beam Elevated 20 Feet Above Ground

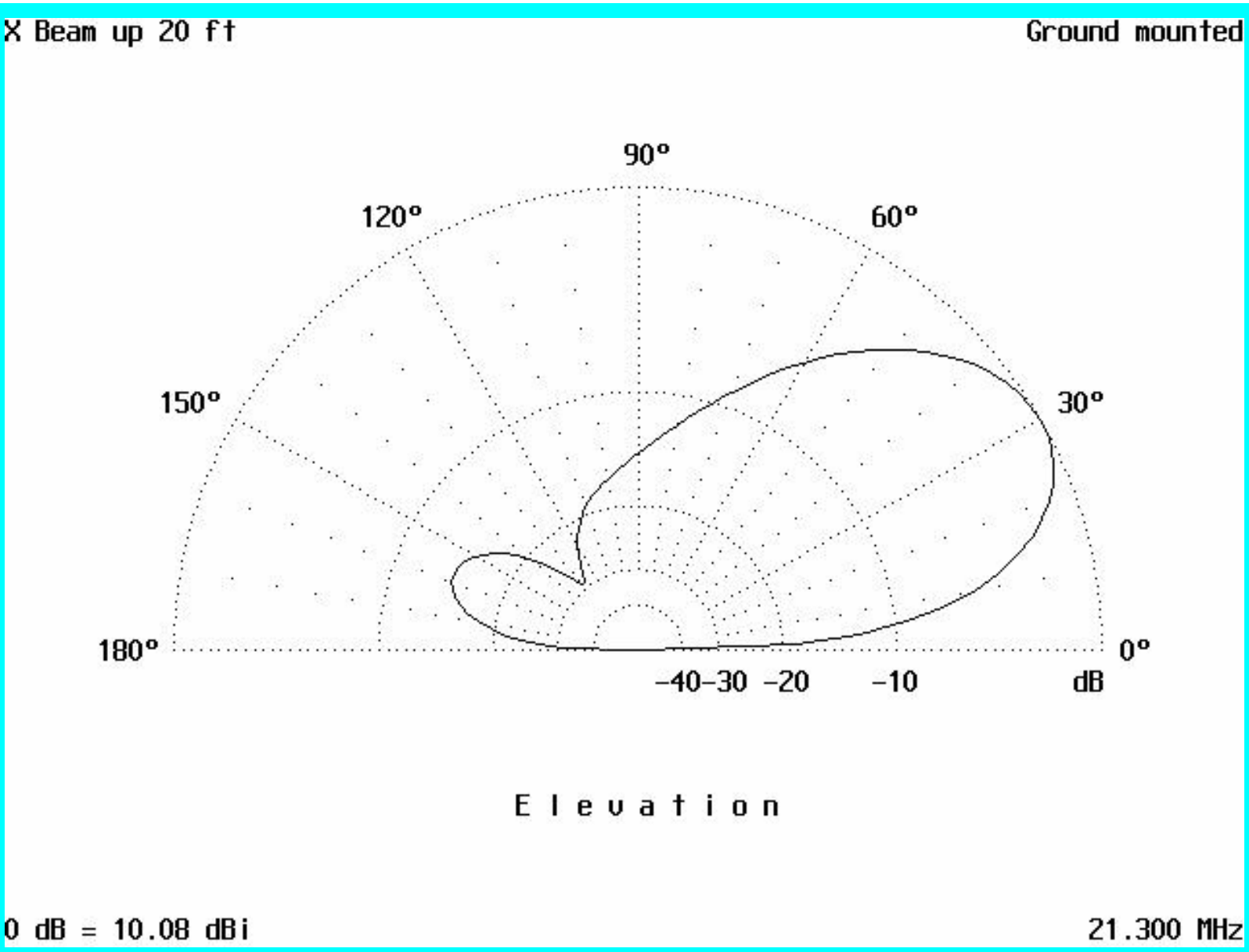


Figure 4
Elevation Pattern of X-Beam Elevated 20 Feet

Note that the gain peaks at a 30 degree launch angle with a peak gain of about 10 dBi or 8 dBd. So this is a fairly good antenna, but not quite as good as the Author indicates. The poor side-lobe level is perhaps the greatest deficiency of the X-Beam, however, it is relatively easy to construct, and portable versions based on telescoping sections of aluminum tubing are easy to transport and erect.

In this case, I'll only consider the design of a 15 meter version, because the 20 meter version requires aluminum spreaders 13' 9" long, and this length is larger than I can construct in my back yard, and the tubing size required is considerably larger than what can be considered as light weight and portable. The spreader (arm) length for 15 meters is only 9' 3" . The director tails are 4' 2", and driven-element tails are 4' 10". The sum of the tails is 9', and

the distance between the tips is sqrt(2) * 9' 3" = 13' 1", which leaves 4' 1" to be spanned by nylon line or cord. You will also need 13' 1" lengths of Nylon cord to span the tips in the front and rear to prevent them from pulling apart when the tails are added. The spreaders (arms) can be assembled from three standard sizes of aluminum tubing beginning with 1/2" OD #22 which has an ID of 0.444 inches. This 1/2" will accommodate 7/16" OD #18 tubing which has an ID of 0.339 inches. The 7/16" tubing will then accommodate 5/16" #20 tubing. I used 4' of 1/2", 4' of 7/16" and 2' of 5/16" tubing for each spreader arm with the first two overlapped by 6" and the outer one overlapped by 4". The hub adds an extra inch from the center of the mast giving a total distance to the tip of exactly 9' 3". The author never states wire size for the tails in the ST article, but in the Antenna Compendium article, he indicates #19 vinyl-covered wire was used. We used #20 vinyl-covered lamp cord by splitting it apart. Various light weight speaker cables work equally well. The directions for tuning this antenna are given in the pair of articles and should be followed in order to get best results. My experience with this antenna mounted about 18' above the ground is that the driving impedance is roughly 25 Ohms when properly tuned. The author indicates a good match with 50 Ohm coax, which probably means that some compromise in performance was accepted to get the good match. The driving impedance does vary with height above the ground.

Parts Required

Discussed above.

References

1. Anderson, Brice, W9PNE, "Designing X-Beams," ARRL Antenna Compendium, Vol. 1, pps. 64-66, 1985.

2. Anderson, Brice, W9PNE, "Horizontal X beams for 15 and 20 Meters," ST, pps. 33-35, March, 1993.

3. Note, all modeling of this antenna has been carried out using Brian Beesley's AO program.

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The Optimized Wide Band Antenna

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15M on 48' Boom
15M on 36' Boom
20M on 48' Boom
10M on 48' Boom
10M on 24' Boom
20M/15M Interlaced
Construction Notes
Conclusion

The Optimized Wideband Antenna

Yagis for 20m - 10m

by Nathan A. Miller
NW3Z
Penn State University

Hello again to every one that read the article I had posted here previously. I have updated all of the previous designs and have added some new ones that you may find interesting. In addition to the 48' boom yagi antennas, I have included a 36' boom 15m antenna as well as a 24' boom 10m antenna that work almost as good as their big brothers. Finally I have posted some pictures of these antennas; both at the K3CR and WP3R contest stations.

Notes regarding stacked installations of antennas:

There were originally two separate versions of the 48' boom 20m antenna: one was designed in free-space and worked well as a stand-alone antenna while the other was designed to operate stacked on a 175' tower. The reason for re-optimization in the stack was that the F/B of the original design was degraded when it was placed into the 6/6/6 stack. To improve the stacked F/B, I unfortunately had to sacrifice the excellent VSWR characteristics of the original antenna. In addition, the F/B of the individual stack-optimized antenna is not as good when it is used independently. The only reason the F/B was degraded is that there is a back-lobe generated at the same elevation angle as the main beam: for the 175' tower this is at 7 degrees. Other than this single lobe, the rear suppression of the stack is quite good. In short, I have been thinking that the small increase in stack performance is not worth the compromises in individual antenna performance.

For this reason, I show only the optimum stand-alone design in this article. I should note that any time antennas are stacked, there is the potential for serious effects on both the F/B and VSWR of the antennas. In general practice, yagis are never stacked closer than .7l. At the K3CR station, the 20m antennas are seperated by .85l which would allow the stand-alone designs to perform relatively well. The 15m antennas are seperated by 1.3l : there will eventually be 3 of the 15m antennas on the 185' tower. While a fourth antenna could have been added without problem, modeling showed that the increase in performance would have been undetectable. For the 10m installation, there will be 4 antennas on the 175' tower with a spacing of by 1.25l. Again, additional antennas could have been added but the performance increase was negligible.

This brings up another question. In the earlier article I posted to WWW.CONTESTING.COM, I mentioned the F/B problems encountered when stacking 6 four element antennas on a large tower. This method will produce a large forward gain but the side-lobe and back suppression will always be limited by the short boom antennas. While there are many advantages to installations such as this, such as the ready availability and low cost of short boom antennas, I still think a smaller number of high performance antennas will produce superior results with less mechanical complexity.

But no matter how good your antennas are, the key to a successful station lies in one simple thing…. RELIABILITY! In stacked installations, the importance of reliable switching and impedance matching systems cannot be over stressed.

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Modeling 6 Long-Boom Yagis

L. B. Cebik, W4RNL

Over the last year or so, I accumulated designs for 5 different 20- meter Yagis with boom between 45 and 55 feet long and having either 5 or 6 elements. After the appearance of a preliminary version of these notes, a 6th was donated. All display an average free space gain from 14.0 to 14.35 MHz of over 10 dBi, with good front-to-back ratios averaging well over 20 dB across the band. It seemed to be an interesting project to model all 6 on the same versions of NEC (in this case, NEC-2 with Leeson corrections and NEC-4, using EZNEC Pro) and do some comparative assessment.

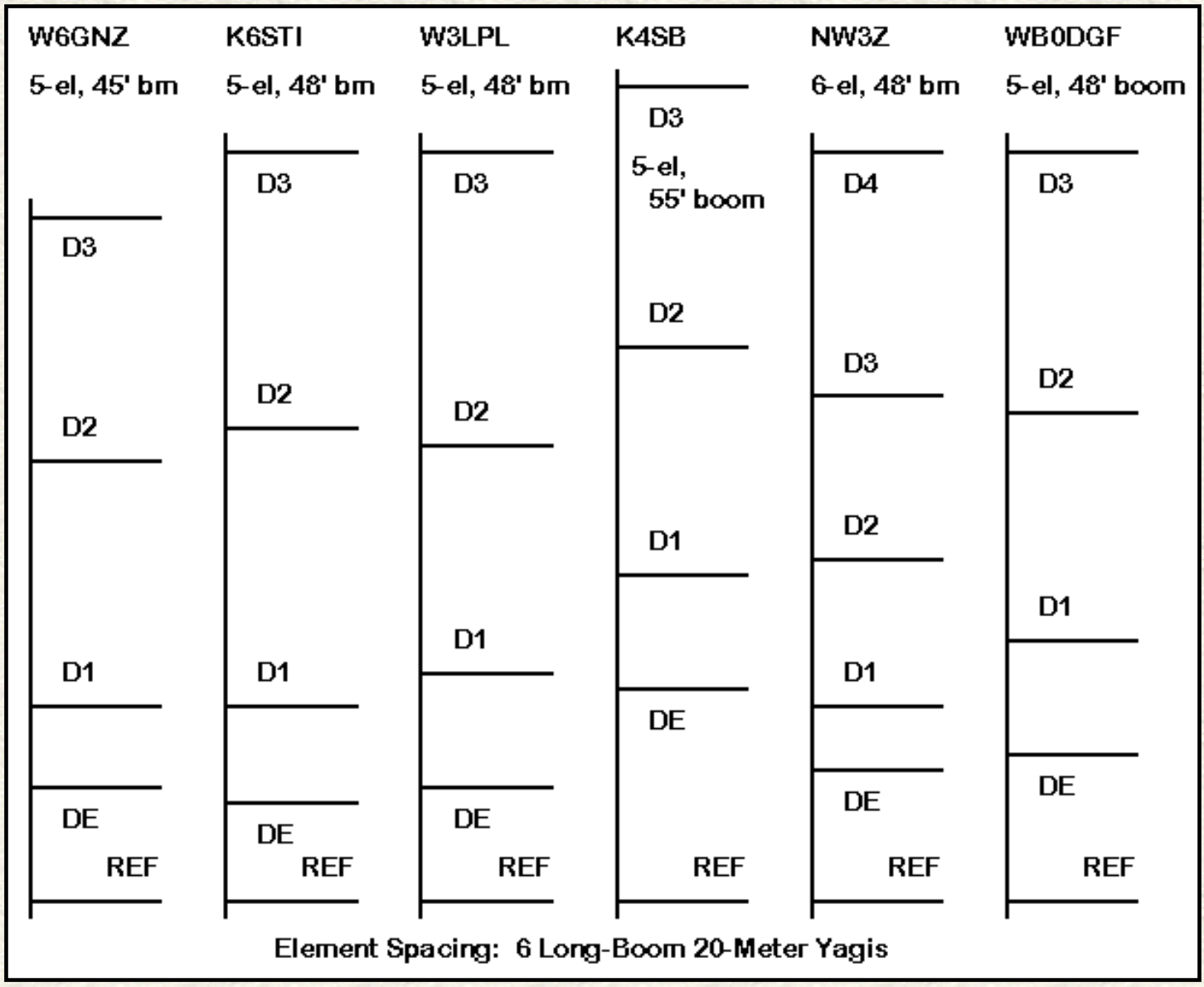
What actually emerged is a lesson in some of the difficulties in making such assessments. Some of those difficulties arise from limitations of the programs (and every version of NEC/MININEC has limitations). Others emerge from differences among models and specification sheets that cannot be resolved without additional data not at hand. Hasty modeling and simplistic conclusions can yield fog where clarity is desired.

The Six Beams

The six beams are the following:

- 1. A 5-element, 45' boom design by W6NGZ, published in *CQ* for October, 1996, p. 22.
- 2. A 5-element, 48' boom design by K6STI, published among the beams in YA, which accompanies the *ARRL Antenna Book*--this is a medium strength model.
- 3. A 5-element, 48' boom design by W3LPL, published on the World Wide Web.
- 4. A 5-element, 55' boom design by K4SB, originally developed in monotaper for 10 meters, modified for this study by me for a somewhat arbitrary element taper (used by permission).
- 5. A 6-element, 48' boom design by NW3Z, published on the World Wide Web.
- 4. A 5-element, 48' sent to me by WB0DGF and once published on the Hy-Gain web site as HG205XLB.

The beams are interesting for their differing design features, not the least of which is the choice in element spacing, as shown in the figure below.



Element Spacing: 6 Long-Boom 20-Meter Yagis

The K6STI has the lowest design feedpoint impedance, which is indicated in the reflector-driver spacing. The W6NGZ design aims for a higher feedpoint impedance and uses a correspondingly wider reflector-driver spacing. The W3LPL design has a comparable feedpoint impedance and hence a similar reflector-driver spacing to the W6NGZ design, despite the differences in boom length. These three designs are similar beyond the driver in that the directors are proportionally spaced, relative to each other. However, do not ignore the ability of D2 to affect feedpoint impedance. The WB0DGF design demonstrates how a different taper schedule may influence the required spacing for a design, since its reflector-drive spacing is greater than the K6STI antenna, but the DGF version has only a moderately higher feedpoint impedance.

The longest model (K4SB) tries for a direct 50-Ohm feed and uses the widest reflector-driver spacing. D1 is not especially different in spacing from the driver relative to the first three models. The truly different design is the NW3Z 6-element Yagi, which uses a moderately wide spacing between the reflector and driver combined with the closely spaced D1 to set the basic parameters of the antenna. However, D2 also plays a crucial role in setting the feedpoint impedance: at correct spacing, the feedpoint impedance is a good coax match; with D2 40" closer to D1, the feedpoint impedance drops to the 25-Ohm neighborhood without seriously affecting the gain and front-to-back characteristics of the antenna.

These notes are not designed to recommend one design over another, either overtly or tacitly. That would require a detailed construction analysis plus the resolution to some modeling questions which I intend to leave open.

Some Modeling Preliminaries

Unfortunately, modeling Yagis within the amateur community has become somewhat of a casual process, without the care and attention to detail used by serious designers, such as the author's of the designs shown here. Whatever the source of the design--YO, Yagi-Max, or simply manual design on AO or EZNEC--evaluation of a design requires good modeling practice.

NEC-2 without a correction factor cannot accurately model a linear element using a tapered-diameter schedule. For that reason, some implementations of NEC-2 include Leeson or similar corrections, which substitute a carefully calculated mono-tapered element for each tapered-diameter element in an array. However, it is possible to disturb the substitute element by paying too little attention to segment length equality. Especially at current nodes (the center of near-1/2 wl elements), segment length needs to be equalized along a linear element element (whether it is a single dipole or a part of a larger array). Since the Leeson corrections modify the length and diameter of a wire but not its segmentation, segment lengths should be equalized in advance. This factor plays an especially important role where the designer may place a very-large diameter short segment at the element center to simulate boom-to-element mounting plates and hardware. Often--as with the K6STI and WB0DGF models--the length of this segment sets the approximatele length of segments in the adjoining sections of the element.

Therefore, for the models at hand, segmentation was increased to yield segment lengths between 8" and 12", depending upon the model. The K6STI and WB0DGF models use the shorter lengths due to their center "lumps." (Note that the WB0DGF model omits the center "lump" on the driven element, while the K6STI model preserves it.) The W6NGZ model uses 12" segments, because they came out most even with the element sections in the design.

Other beams used 8" (W3LPL) or 10" (K4SB, NW3Z) per segment for similar reasons: the chosen segment length turned out to be the best way to achieve roughly equal length segments on the most elements. Despite best efforts, some variations inevitably result. Designers will always have, for any segment length chosen, a section that comes out to be n.5 segments long. Some will choose center sections that for any reasonable segment length seem to require an even number of segments. Adding extra segment is necessary for the driver--and hence, for the other elements in order to keep segment junctions as well aligned as possible.

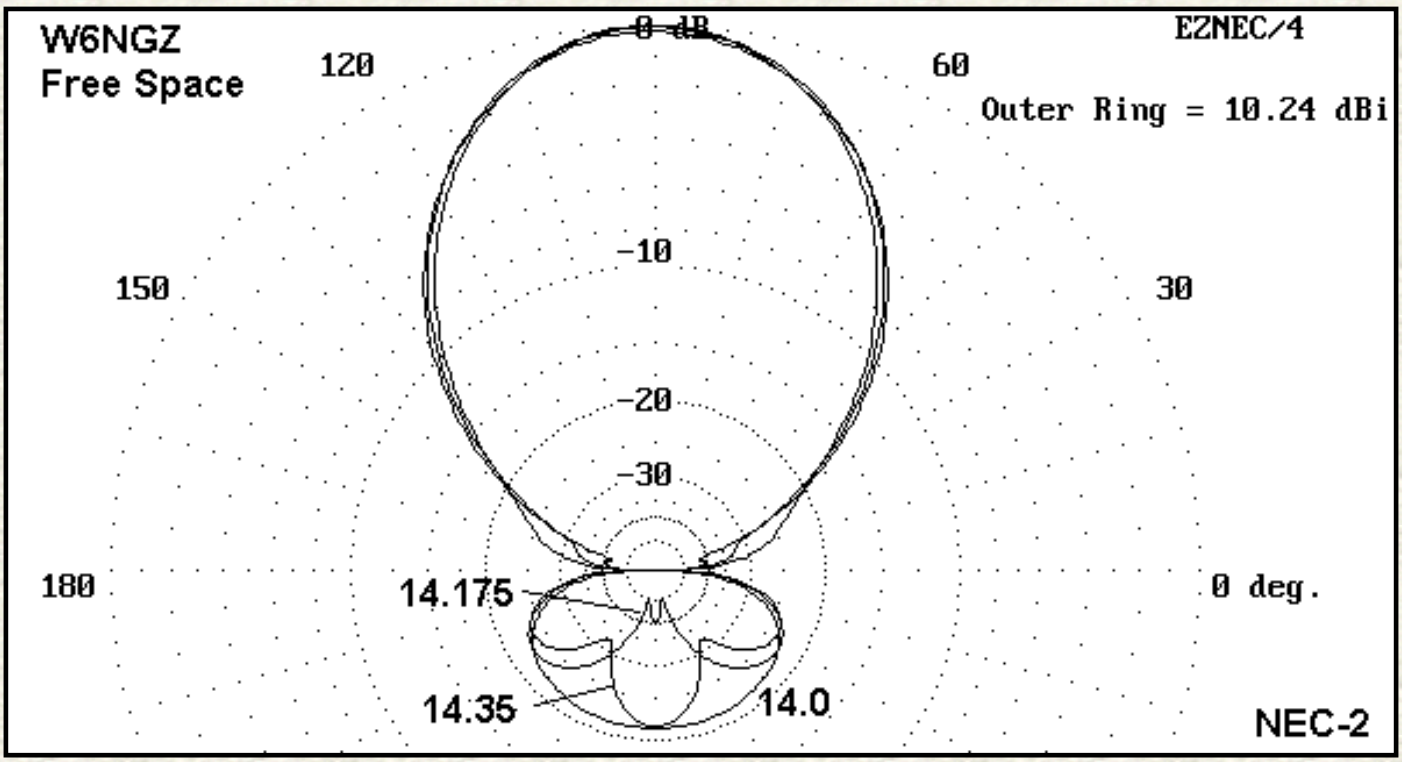
NEC-4, by its initial reports, overcomes the NEC-2 weakness with tapered-diameter elements. Unfortunately, the initial reports have proven overly optimistic. Although NEC-4 is a vast improvement over uncorrected NEC-2, it falls short of the standard set by NEC-2-corrected, which also correlates with MININEC results. (MININEC does not have a tapered-diameter element problem, although it has other limitations.)

The NEC-4 limitation is somewhat minor for general purpose modeling, unless the element shows too large a diameter change between wires in the element, especially close to the current node, with the driven element the most crucial. This condition exists with the K6STI model.

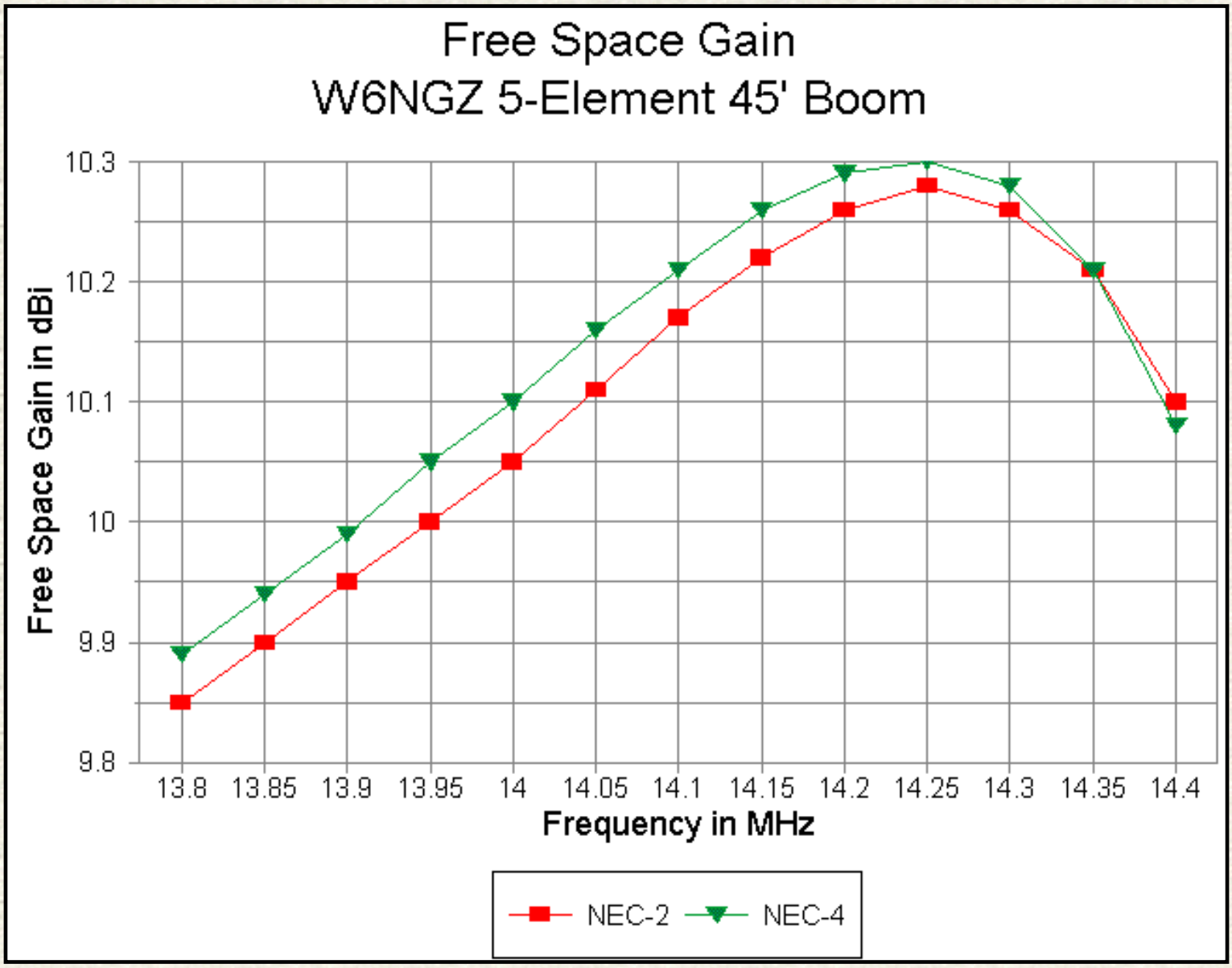
In order to show the modeling differences between NEC-2-corrected and NEC-4, I have modeled all 6 beams in both programs via EZNEC Pro. (I could have applied the Leeson equations to the NEC-4 runs, but that would have produced the same results as the NEC-2 runs.) I graphed data comparatively for NEC-2-corrected and for NEC-4, to include free space gain in dBi, 180-degree front-to-back ratio, feedpoint resistance, and feedpoint reactance. SWR data is also shown wherever the source impedance is close enough to 50-Ohms to make such data relevant. All graphs run from 13.8 to 14.4 MHz in order to show downward slides of NEC-4 curves.

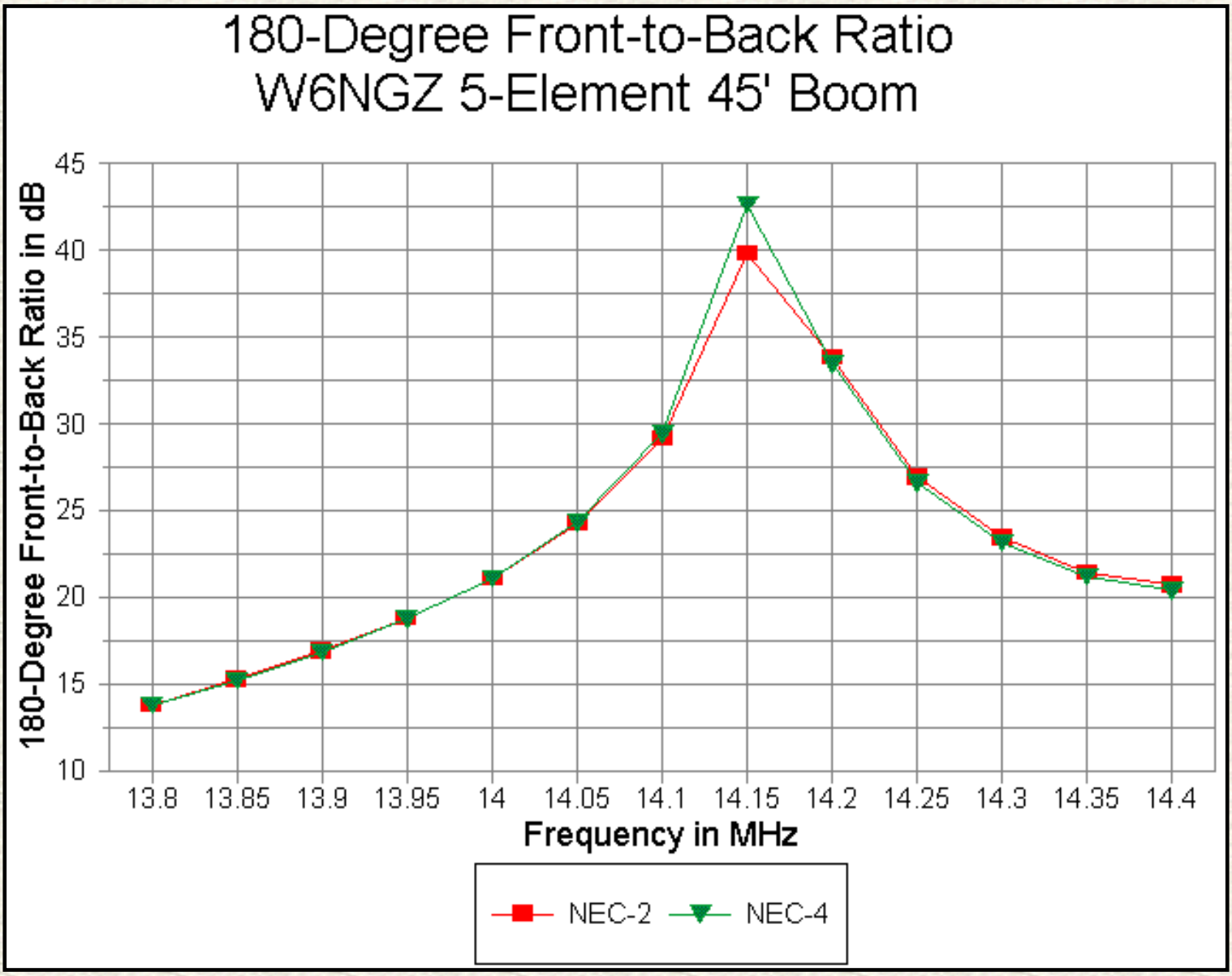
Let's look at the designs individually.

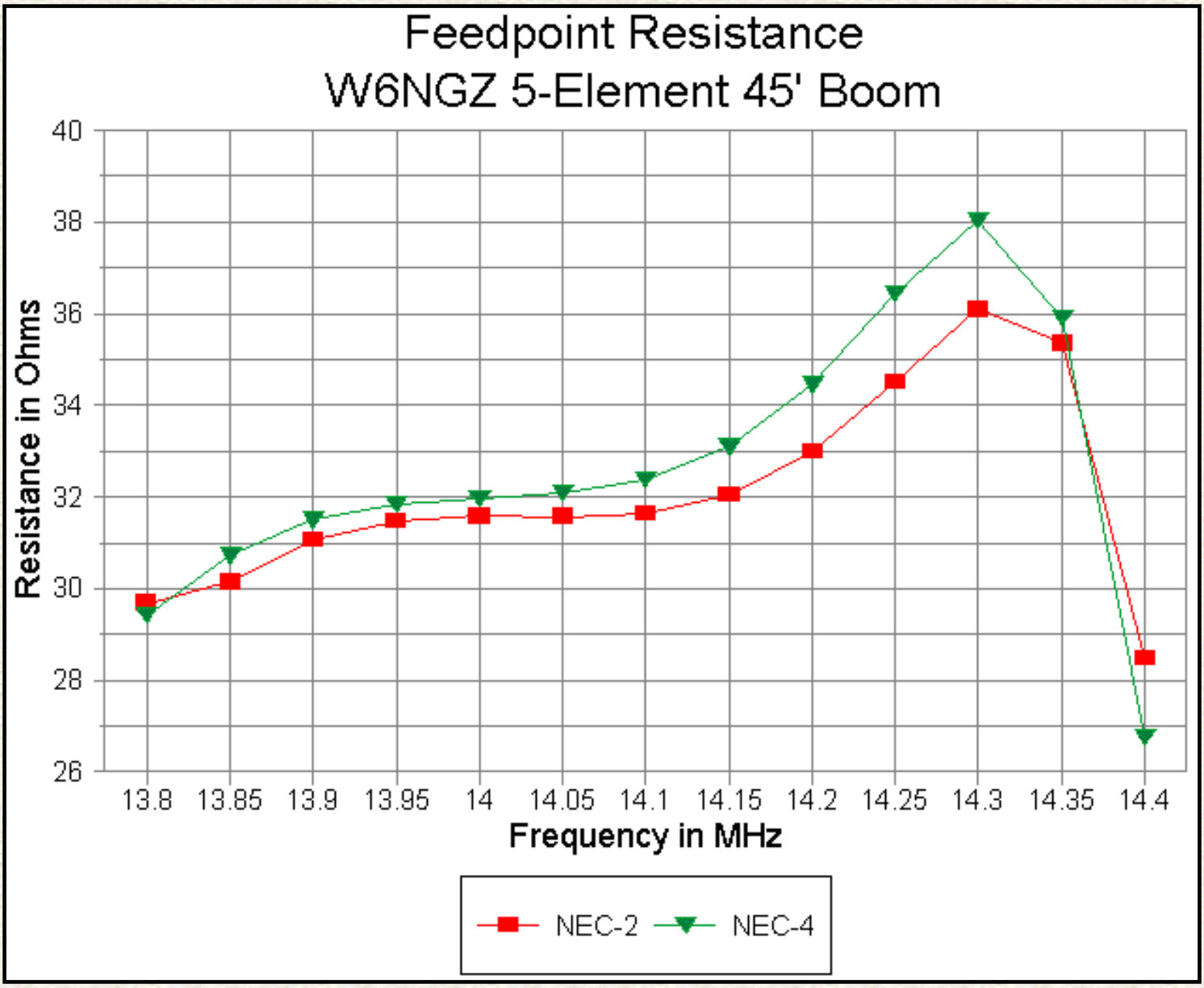
The W6NGZ 5-element, 45' boom model

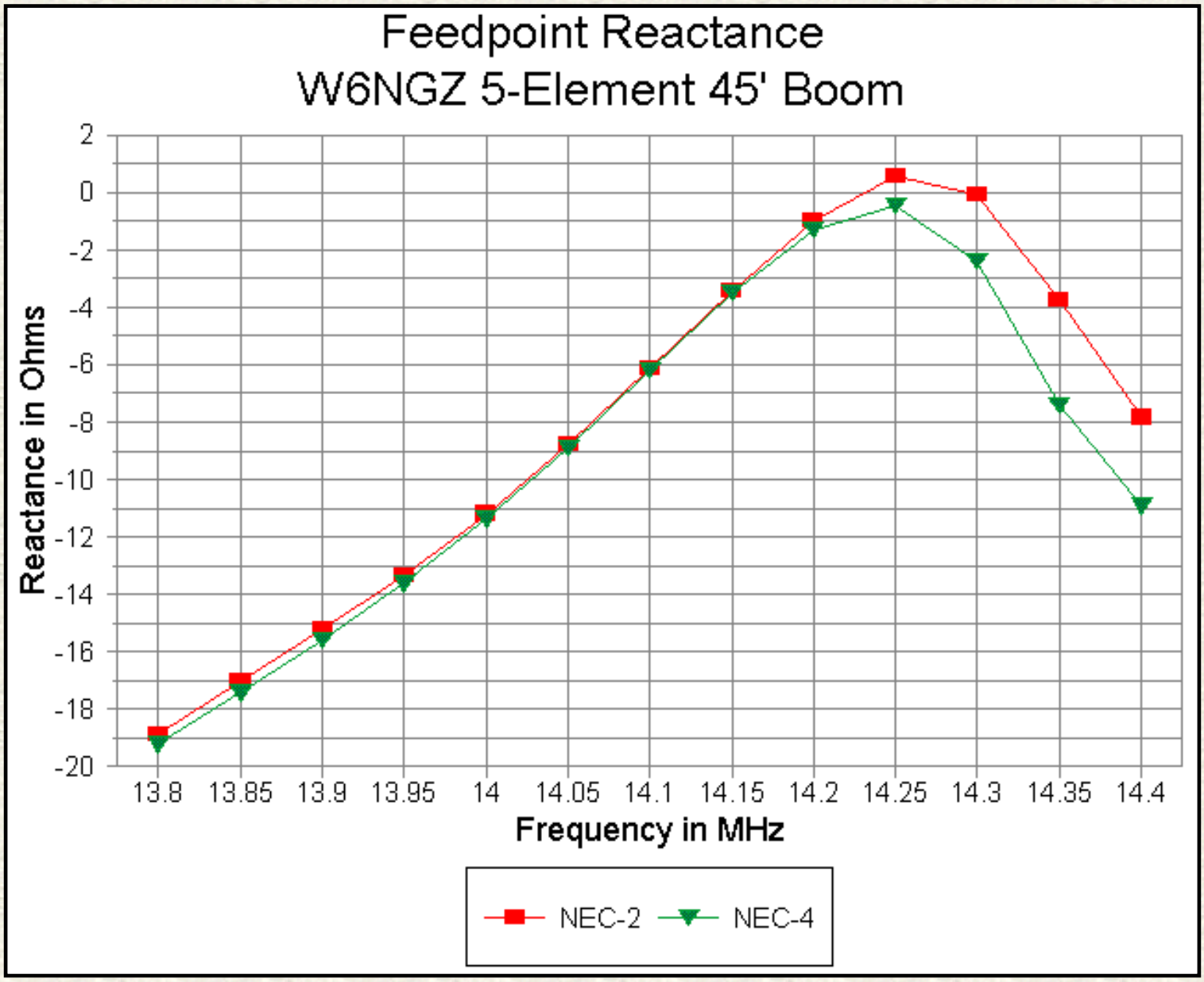


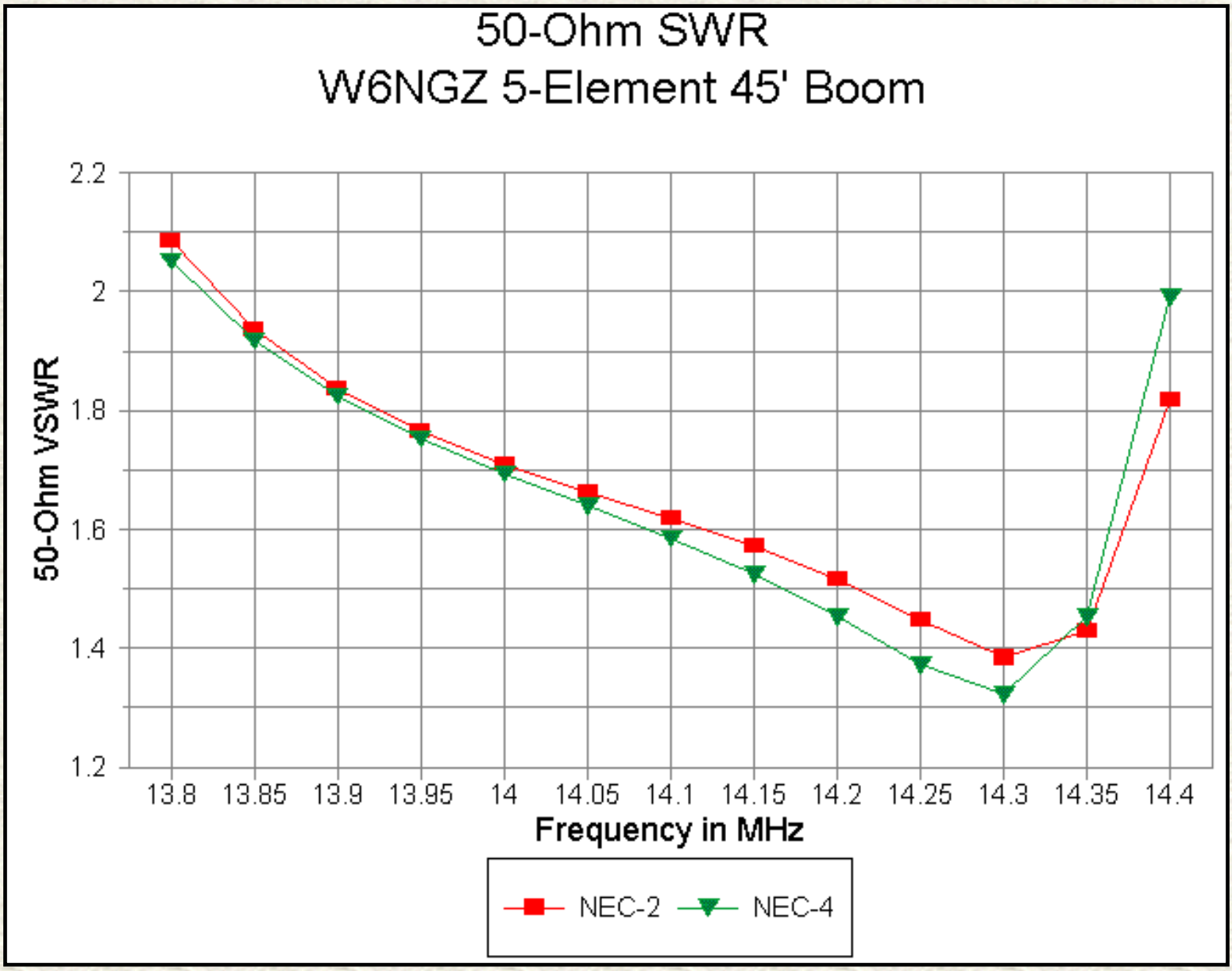
The W6NGZ model is the shortest of the five, but it sustains a free space gain of over 10 dBi across the band, with better than 20 dB front-to-back ratio everywhere on 20. The element taper is from the *CQ* article, and it contains no compensation for element mounting plates at the element center. Within these constraints, NEC-4 and NEC-2-corrected tell very similar graphical stories about the beam. Differentials are marginal at best, although a wee bit of excess gain estimation and a slide of the curves downward in frequency are both evident.


Free Space Gain
W6NGZ 5-Element 45' Boom
Free Space Gain in dBi
10.3
10.2
10.1
10
9.9
9.8
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4


180-Degree Front-to-Back Ratio
W6NGZ 5-Element 45' Boom
180-Degree Front-to-Back Ratio in dB
45
40
35
30
25
20
15
10
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4

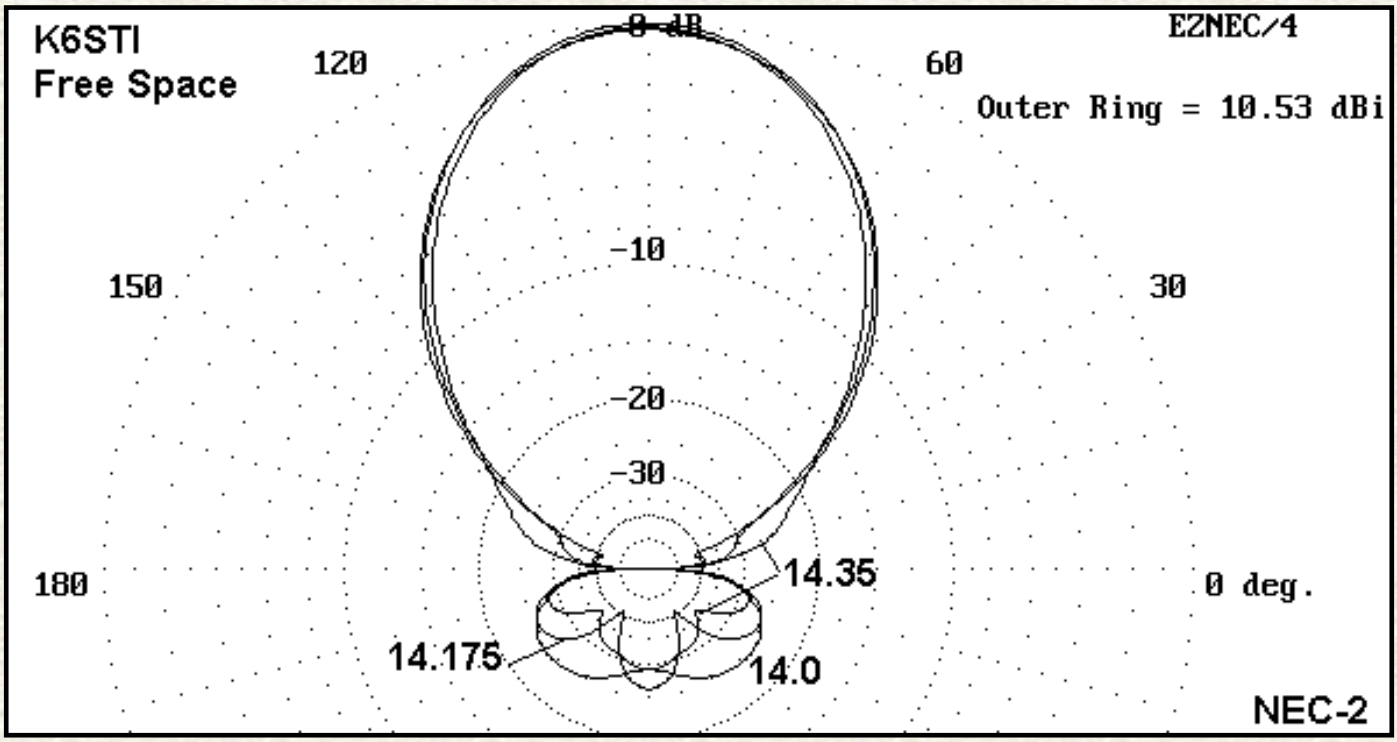

Feedpoint Resistance
W6NGZ 5-Element 45' Boom
Resistance in Ohms
40
38
36
34
32
30
28
26
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4


Feedpoint Reactance
W6NGZ 5-Element 45' Boom
Reactance in Ohms
2
0
-2
-4
-6
-8
-10
-12
-14
-16
-18
-20
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4

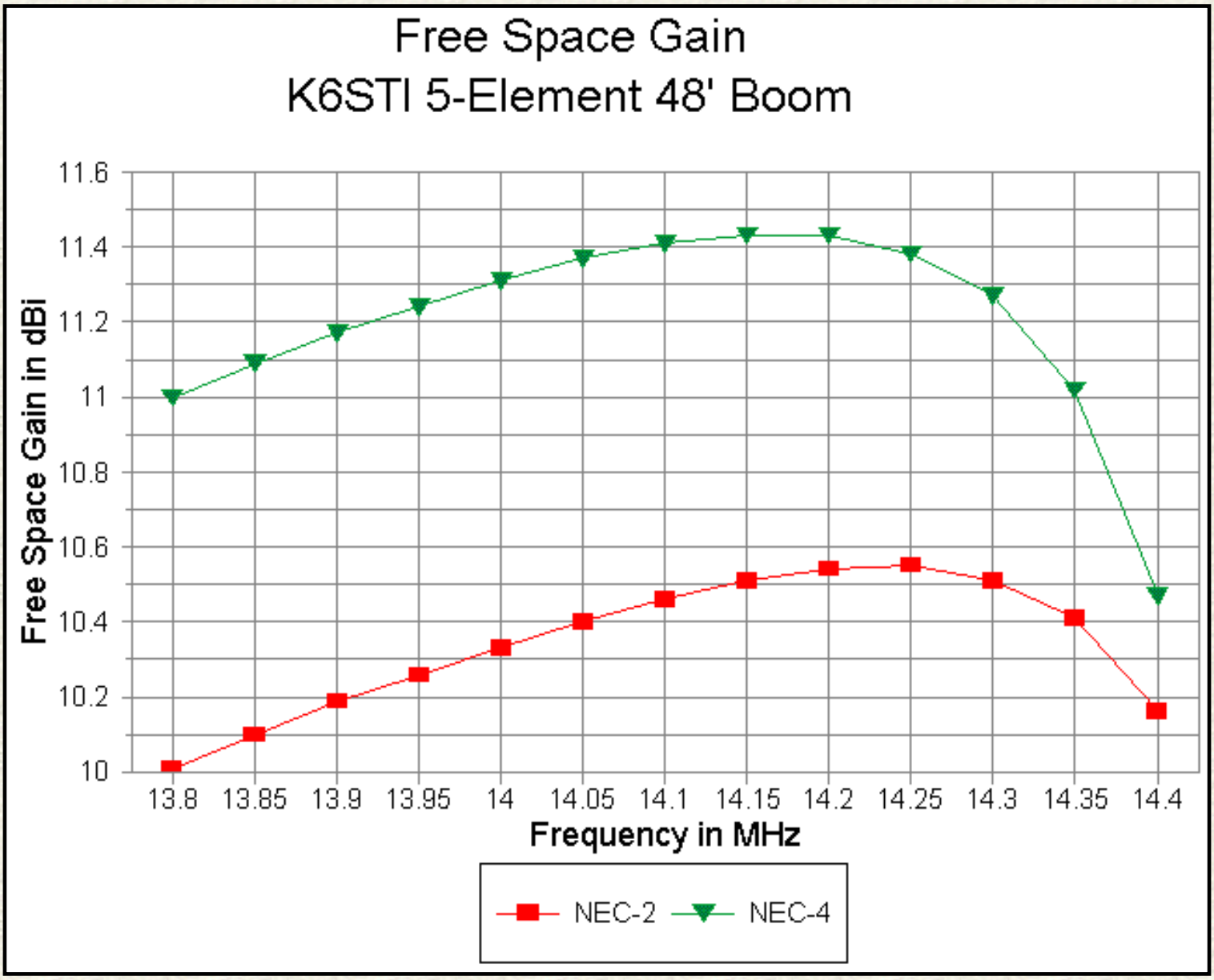


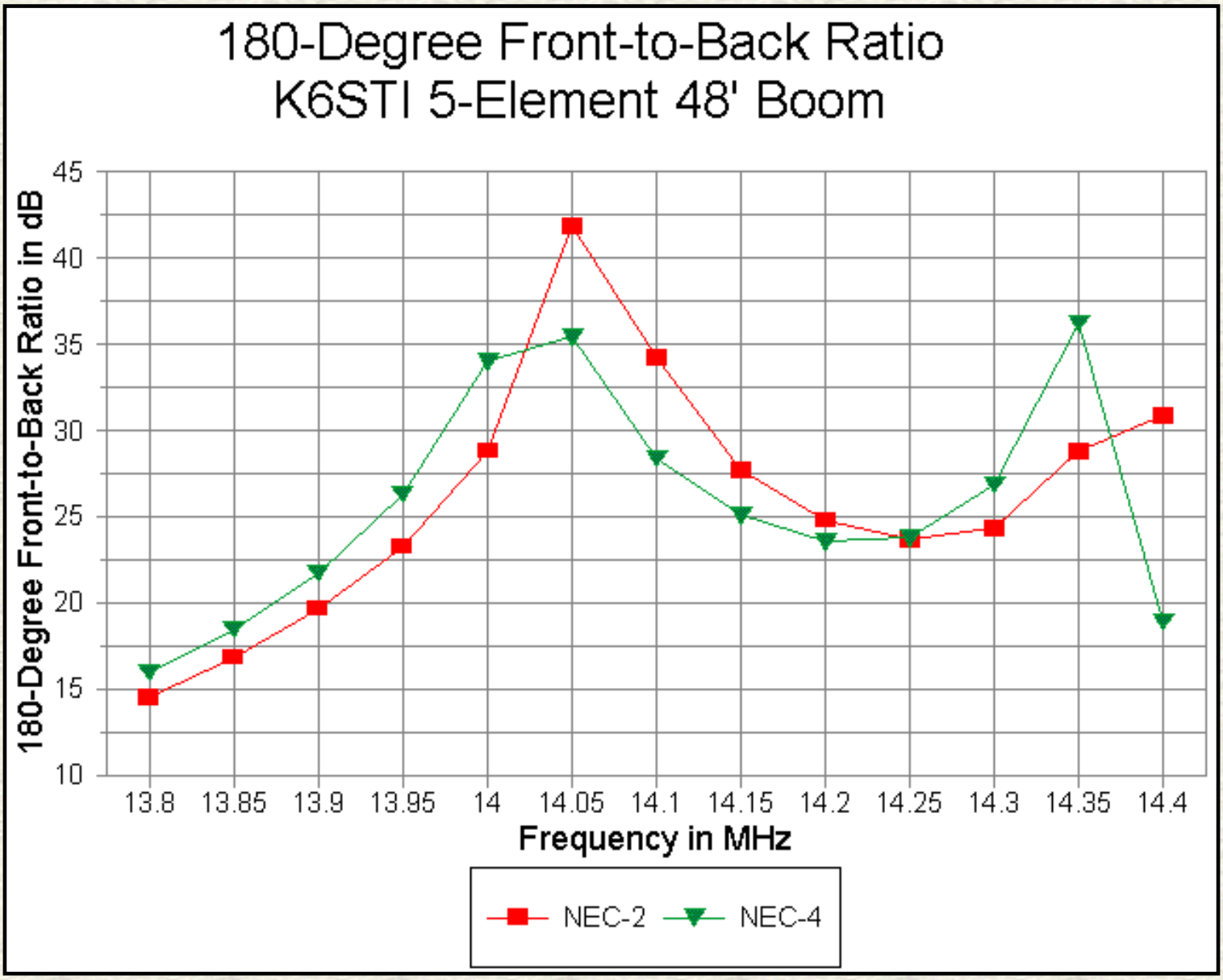
As the graphics make plain, the antenna close being a very good match for 50-ohm cable.

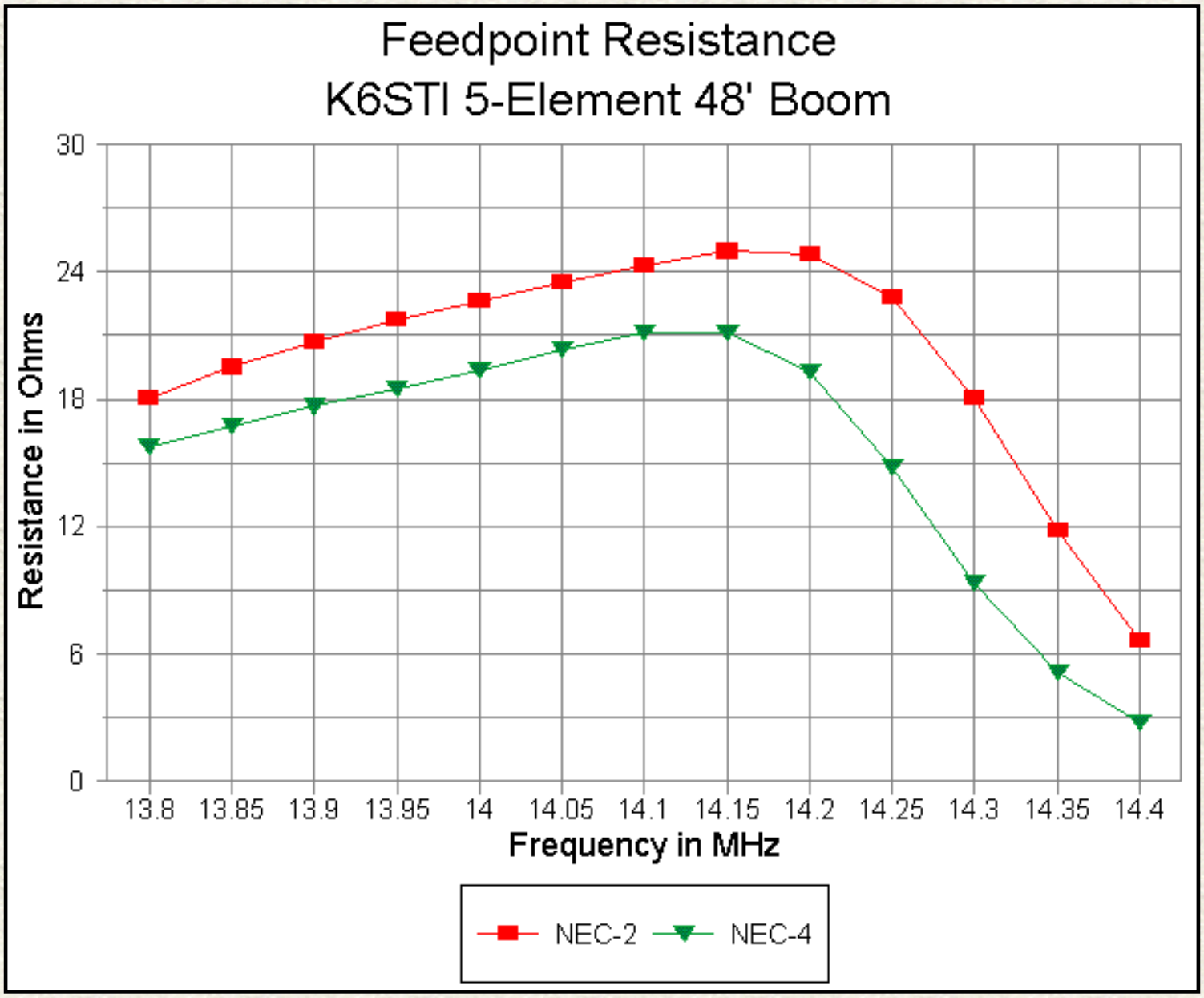
The K6STI 5-element, 48' boom model

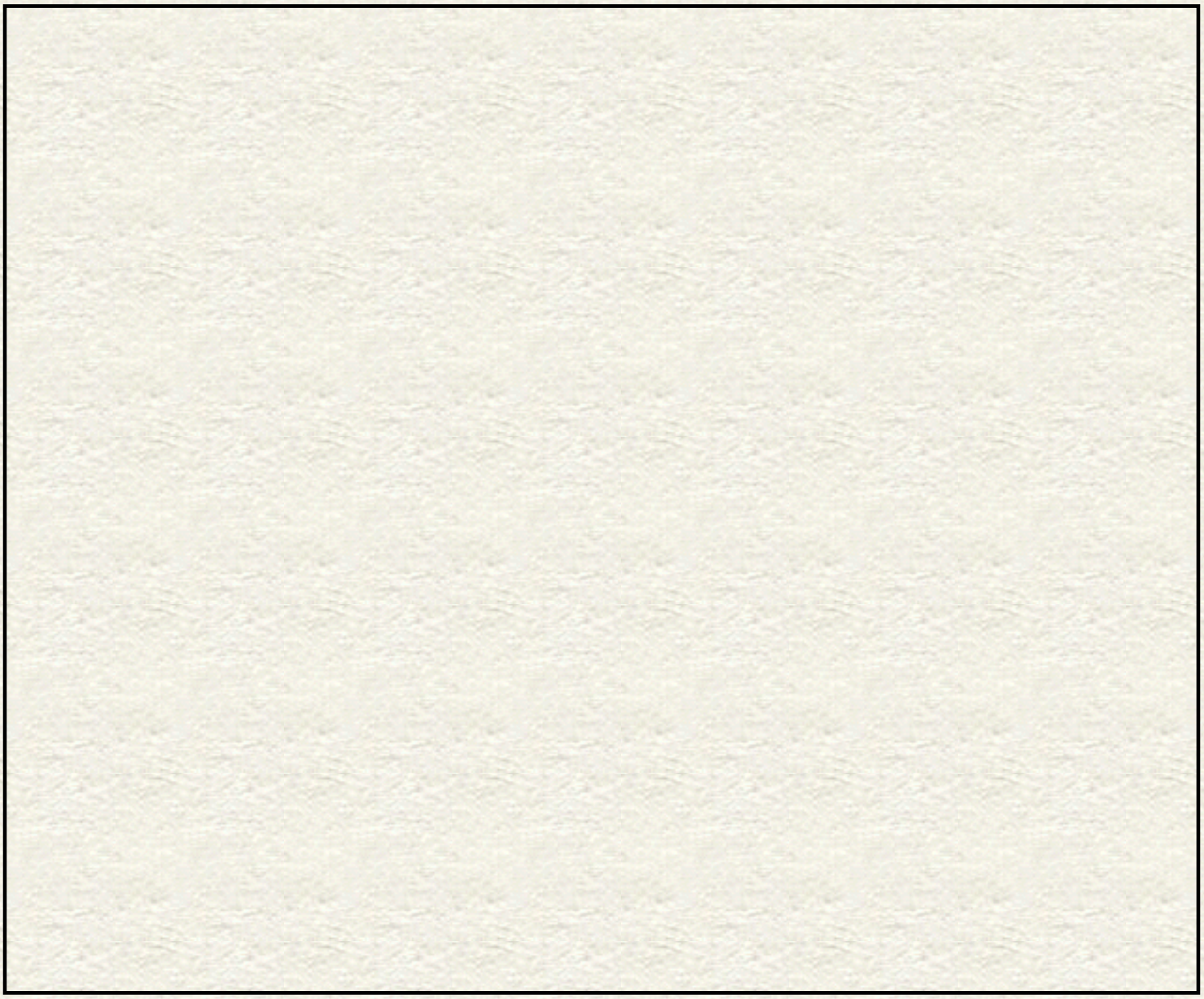


The K6STI design, with its center "bulge" in place, models quite well in NEC-2-corrected. It also displays very well the weakness of NEC-4 in dealing with the large change of diameter so close to the source or highest current point on the elements, despite extensive efforts to equalize segment lengths. NEC-4 gain data is almost a full dB higher than the more reasonable NEC-2 data, with a curve displacement lower in frequency, which also shows up on the front-to-back graph. The source resistance and reactance graphs show a similar slide, with the resistance values low in the NEC-4 run. Since current, impedance, and far field strength are related, the coincidence of these value differences from those in the NEC-2 run are not unexpected.


Free Space Gain
K6STI 5-Element 48' Boom
Free Space Gain in dBi
11.6
11.4
11.2
11
10.8
10.6
10.4
10.2
10
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4

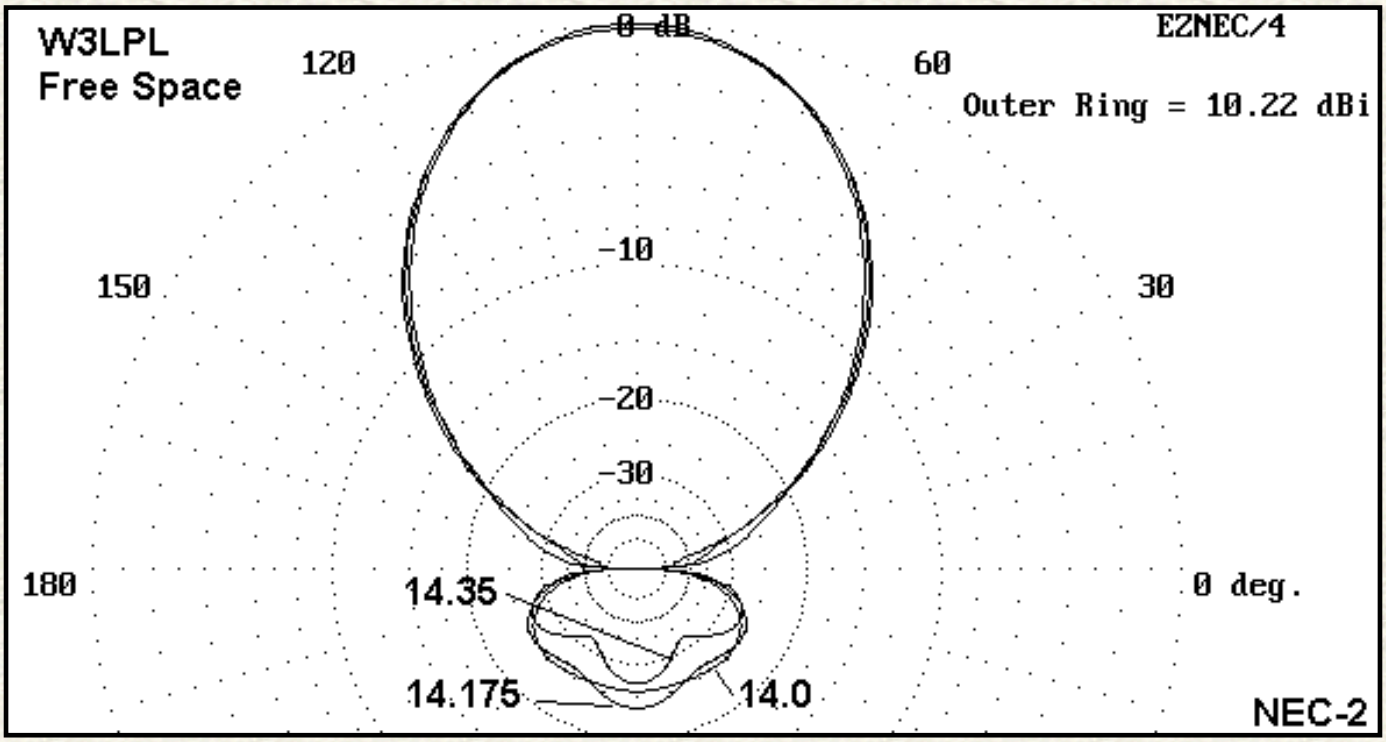

180-Degree Front-to-Back Ratio
K6STI 5-Element 48' Boom
180-Degree Front-to-Back Ratio in dB
45
40
35
30
25
20
15
10
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4


Feedpoint Resistance
K6STI 5-Element 48' Boom
Resistance in Ohms
30
24
18
12
6
0
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4

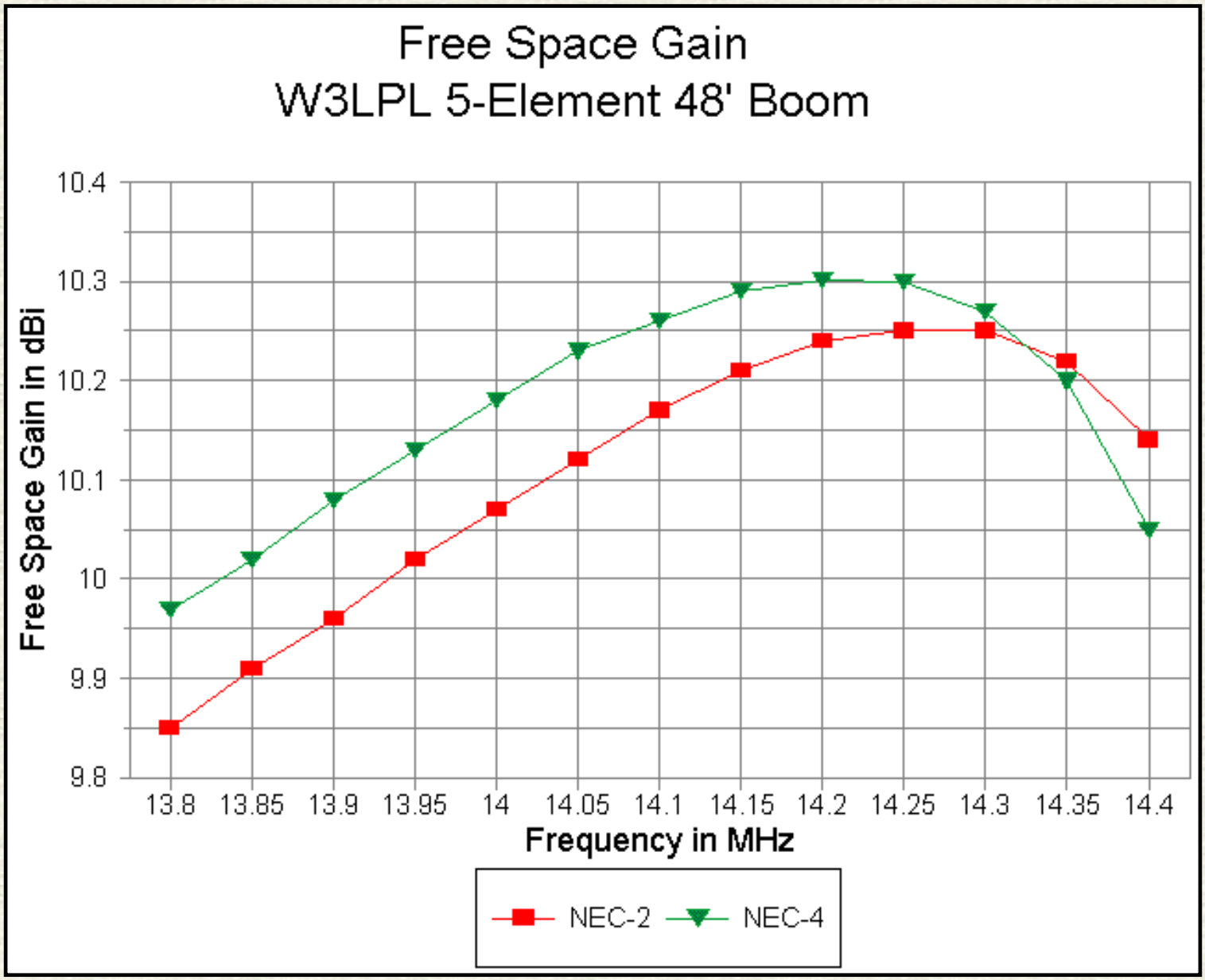


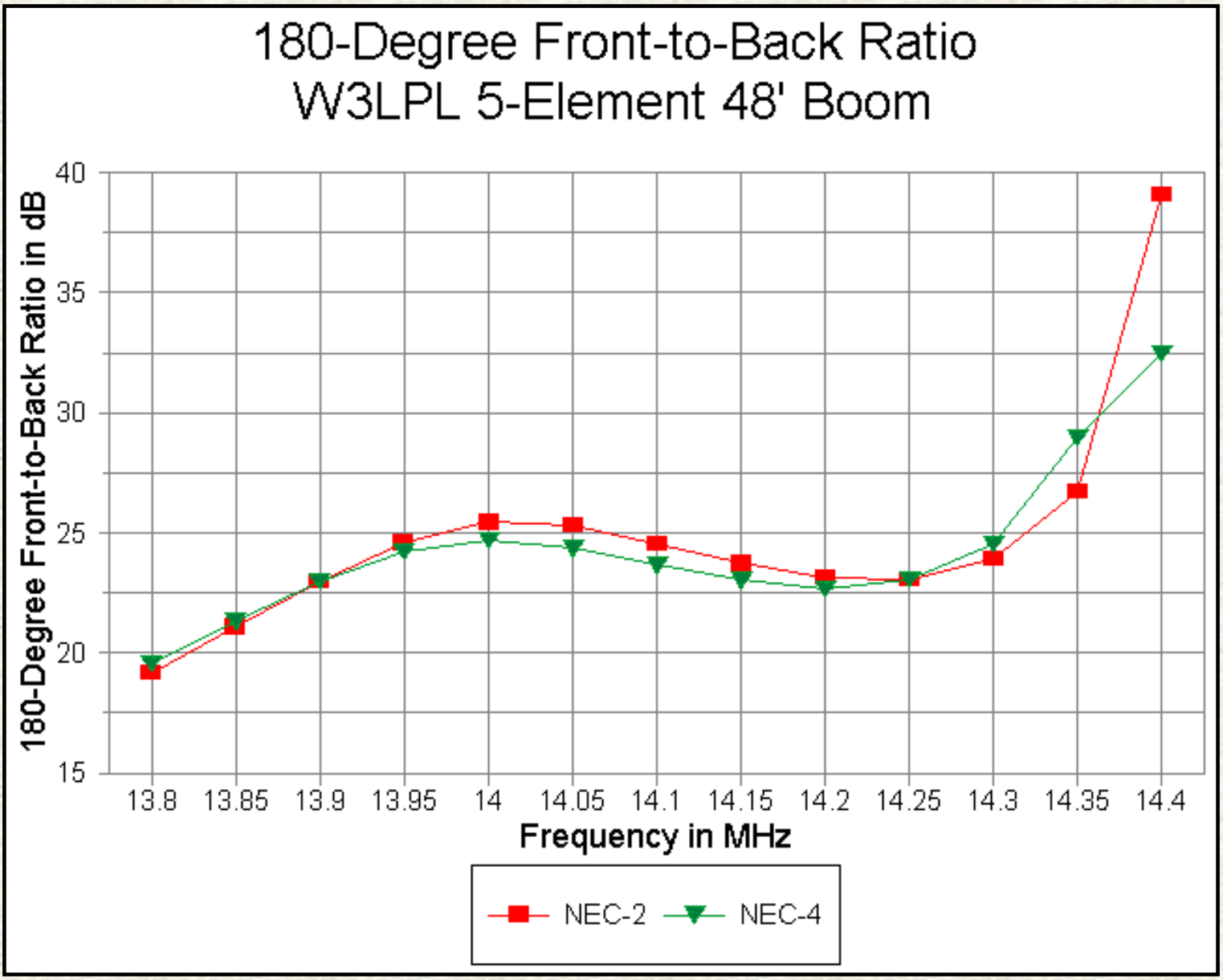
No SWR curve is shown, since the antenna is designed for a matching system such as a beta/hairpin match.

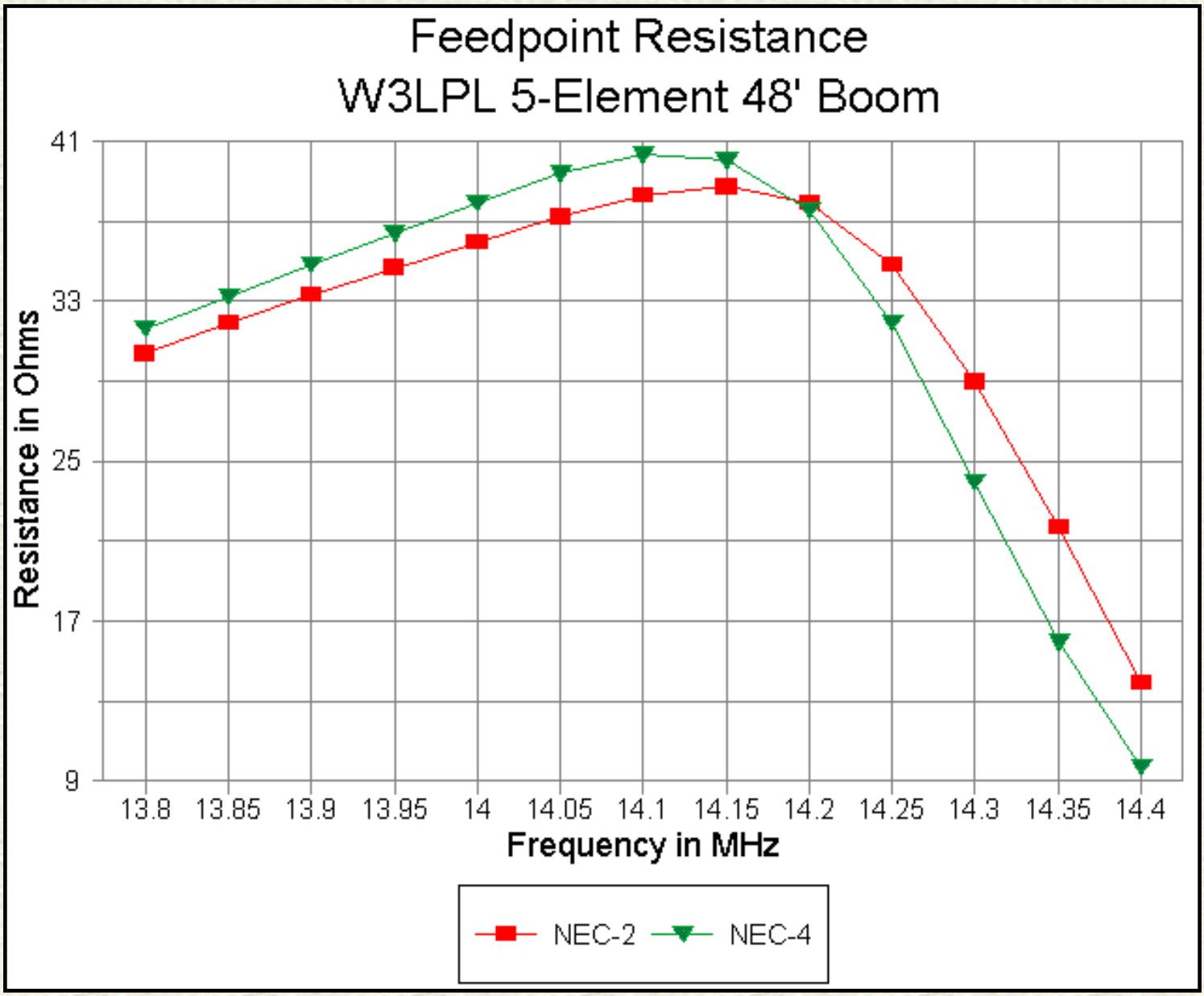
The W3LPL 5-element, 48' boom model

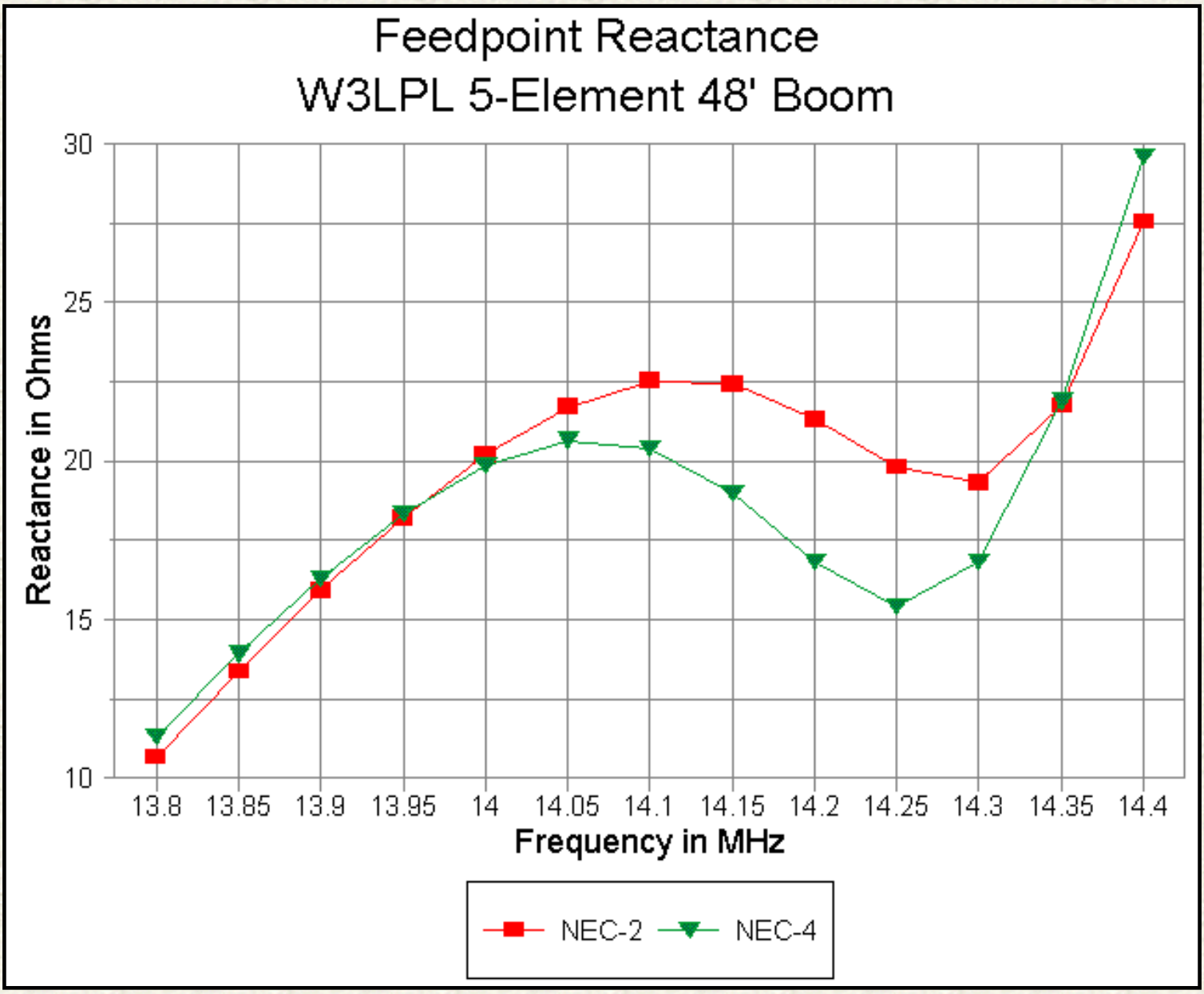


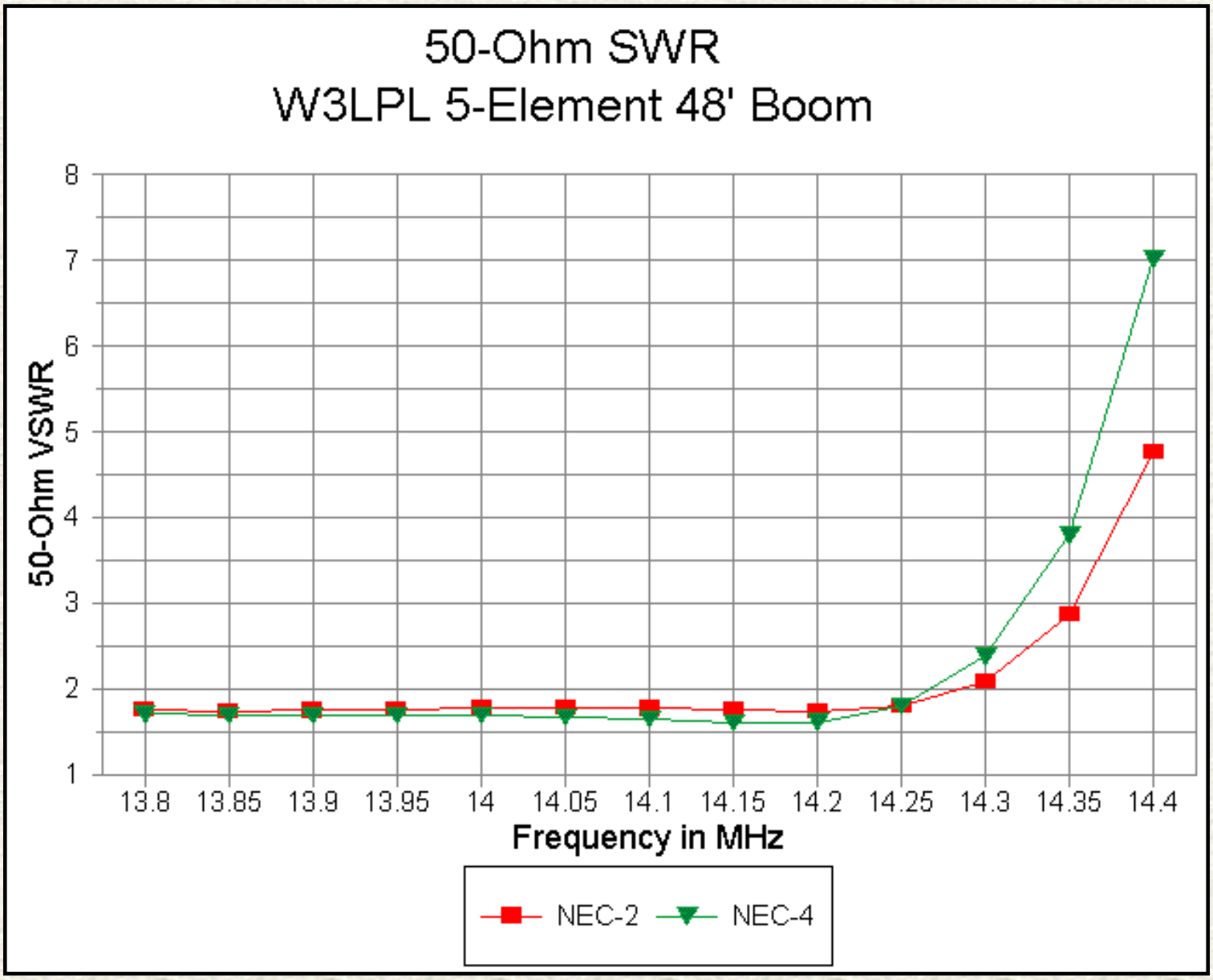
The W3LPL beam was designed for use with a Tee match, which neither NEC-2 nor NEC-4 can physically model accurately. Interestingly, W3LPL designed the antenna with non-conductive boom-to-element mounting plates in mind. Nonetheless, the curves--plus a look at the antenna characteristics just below 14 MHz--suggests that there is an approximate 50 kHz offset in frequency, with the elements in the model again playing short in the NEC-4 run, relative to the NEC-2-corrected run. Gain is still a little high in NEC-4. Given the W6NGZ model runs, this much difference is somewhat unexpected. The chief difference in the modeled data is that the W3LPL antenna shows considerable and constant inductive reactance, due to its intended use with a Tee-matching system.


Free Space Gain
W3LPL 5-Element 48' Boom
Free Space Gain in dBi
10.4
10.3
10.2
10.1
10
9.9
9.8
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4


180-Degree Front-to-Back Ratio
W3LPL 5-Element 48' Boom
180-Degree Front-to-Back Ratio in dB
40
35
30
25
20
15
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4

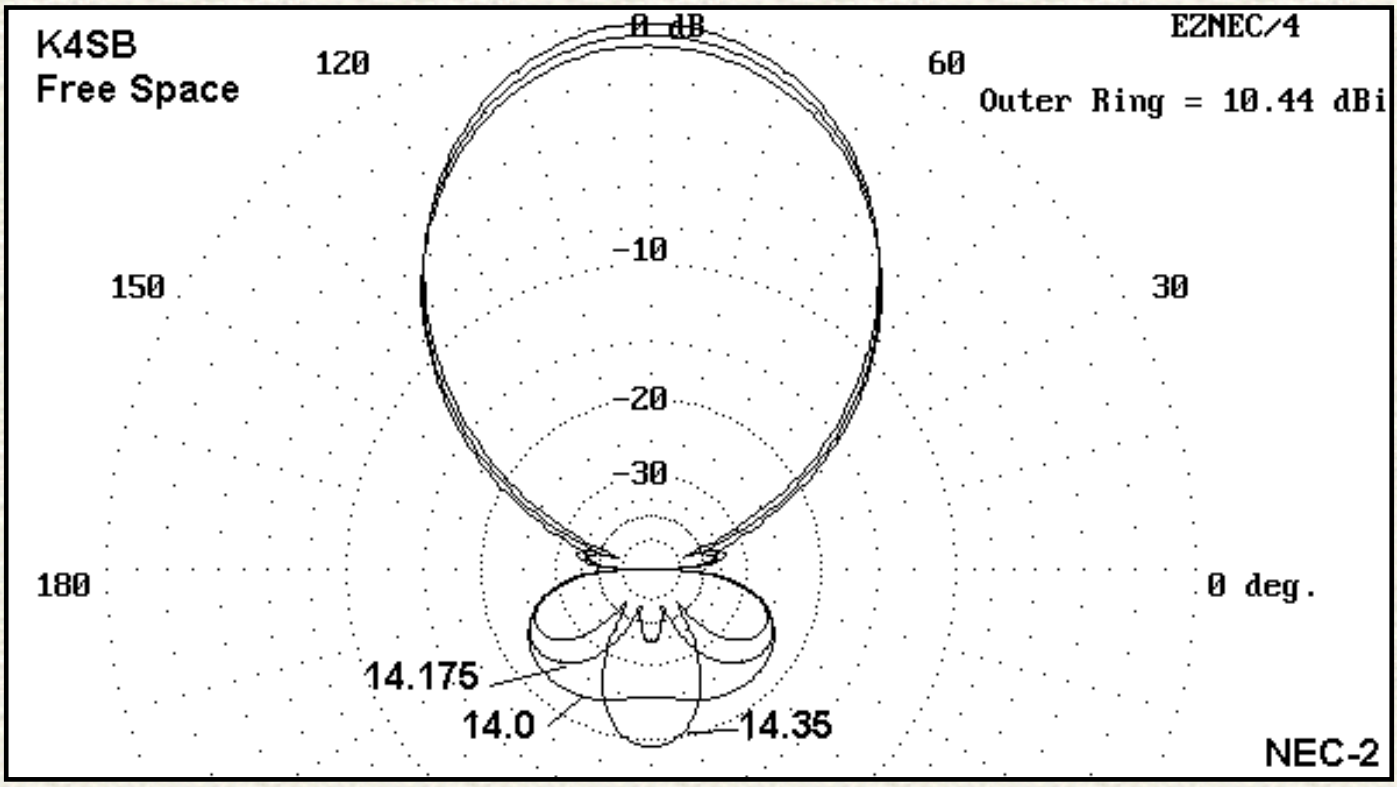

Feedpoint Resistance
W3LPL 5-Element 48' Boom
Resistance in Ohms
41
33
25
17
9
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4


Feedpoint Reactance
W3LPL 5-Element 48' Boom
Reactance in Ohms
30
25
20
15
10
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4



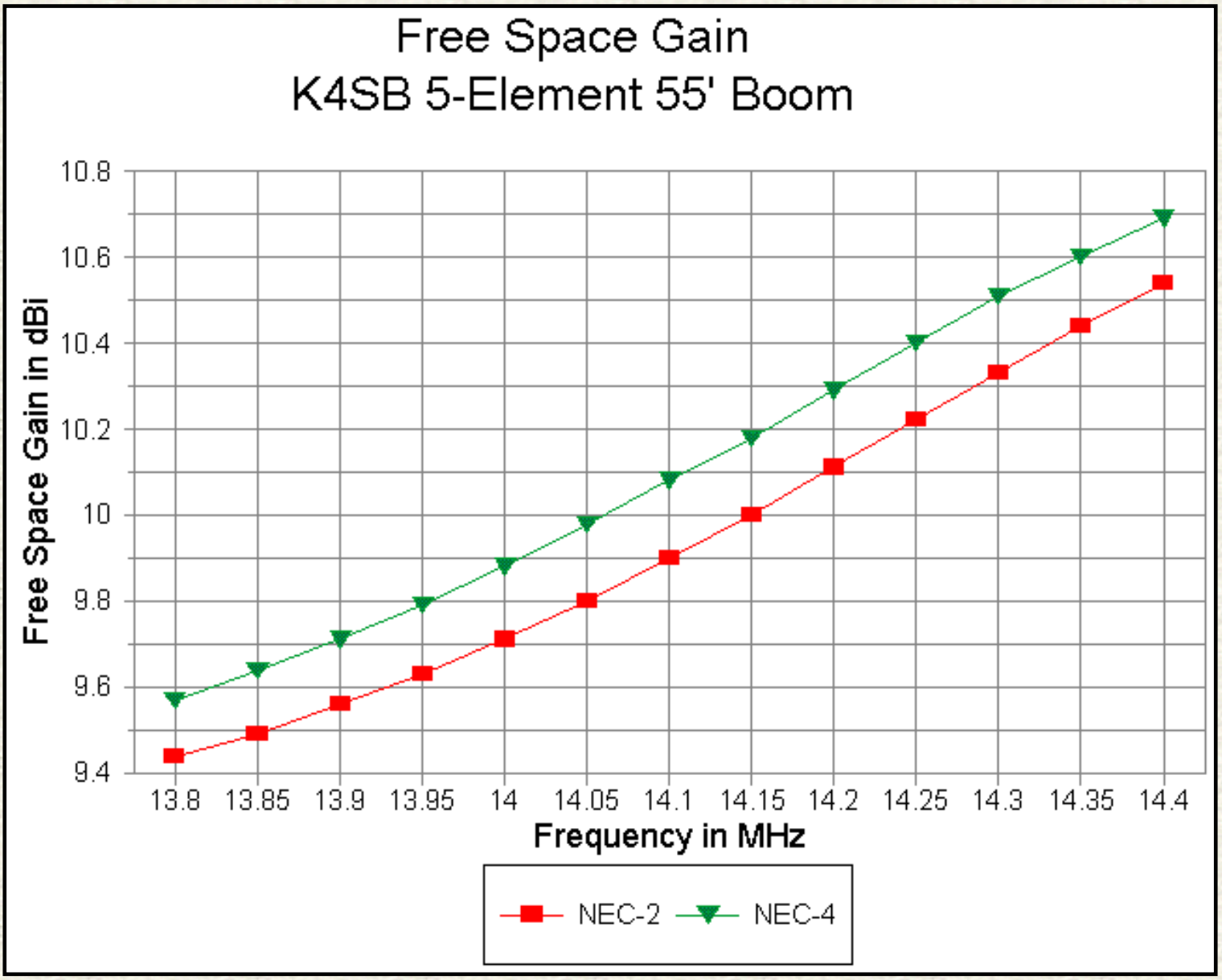
I have included the 50-Ohm SWR curve for the W3LPL antenna because it makes a good contrast with that of the W6GNZ antenna, even though the antenna designer specifically introduced the inductive reactance in the driver to set up conditions for a no-capacitor Tee match system for a 200-Ohm feed impedance. Shortening the driver to resonance in an attempt to reduce the SWR will result in a decrease in the resistive component so that no great improvement actually occurs. However, those wishing a direct 50-Ohm feed without a Tee match may wish to consider a beta match using a shunt capacitor, instead of the usual shunt inductor which is applied when the driver element is capacitively reactive. A capacitor value of about 90 Ohms reactance (about 125 pF) with sufficient voltage and current handling capabilities would come close to serving the need.

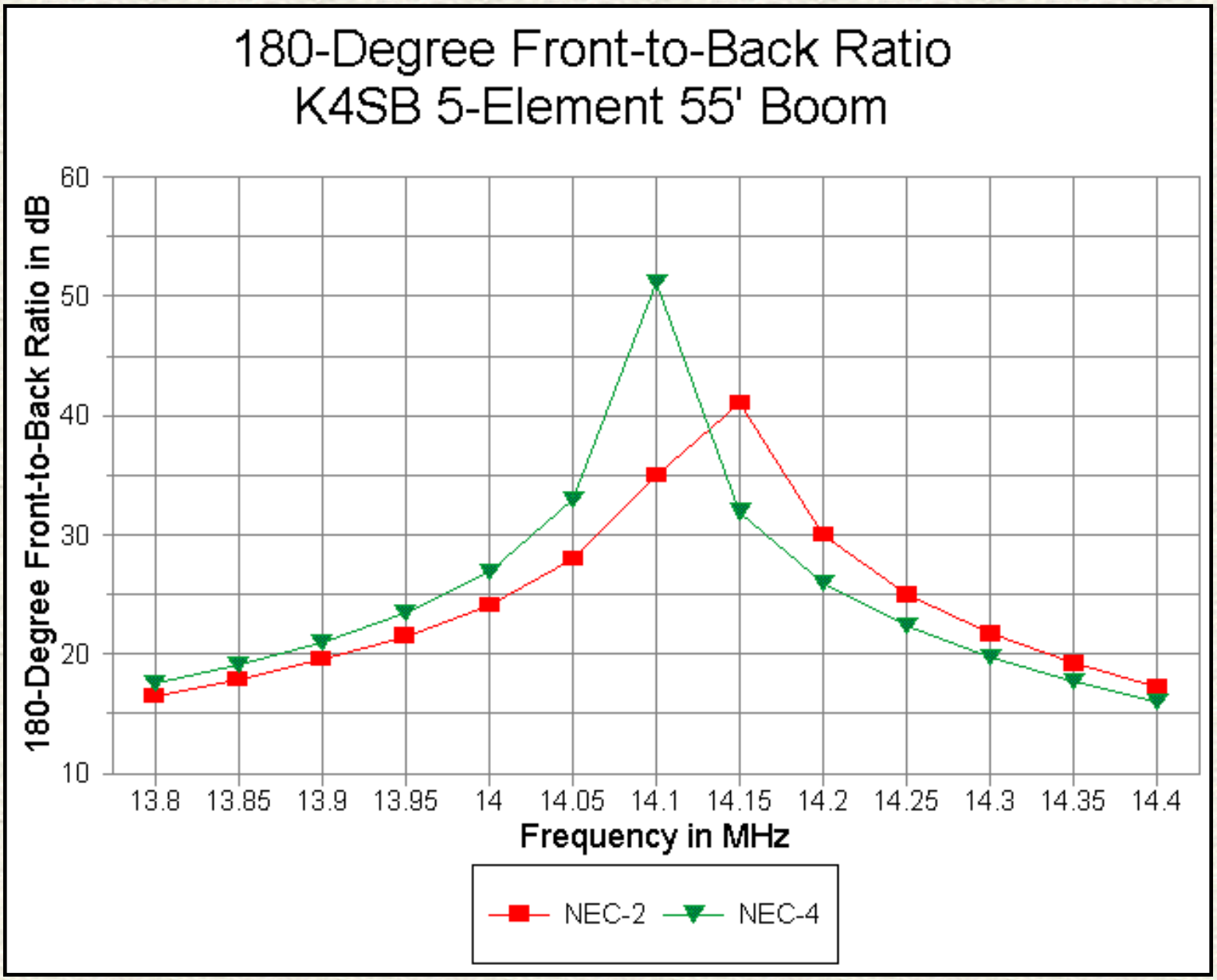
The K4SB 5-element, 55' boom model

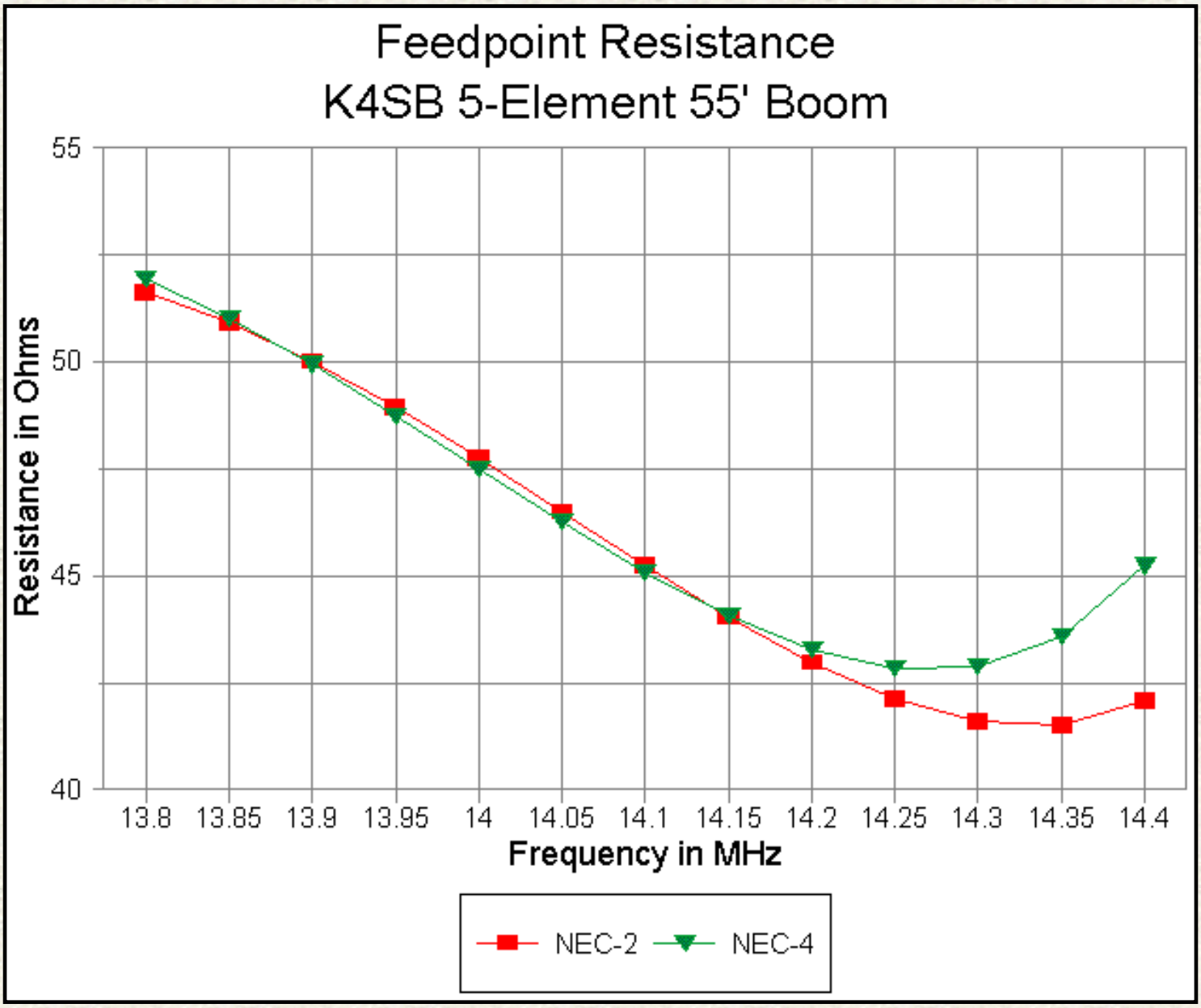


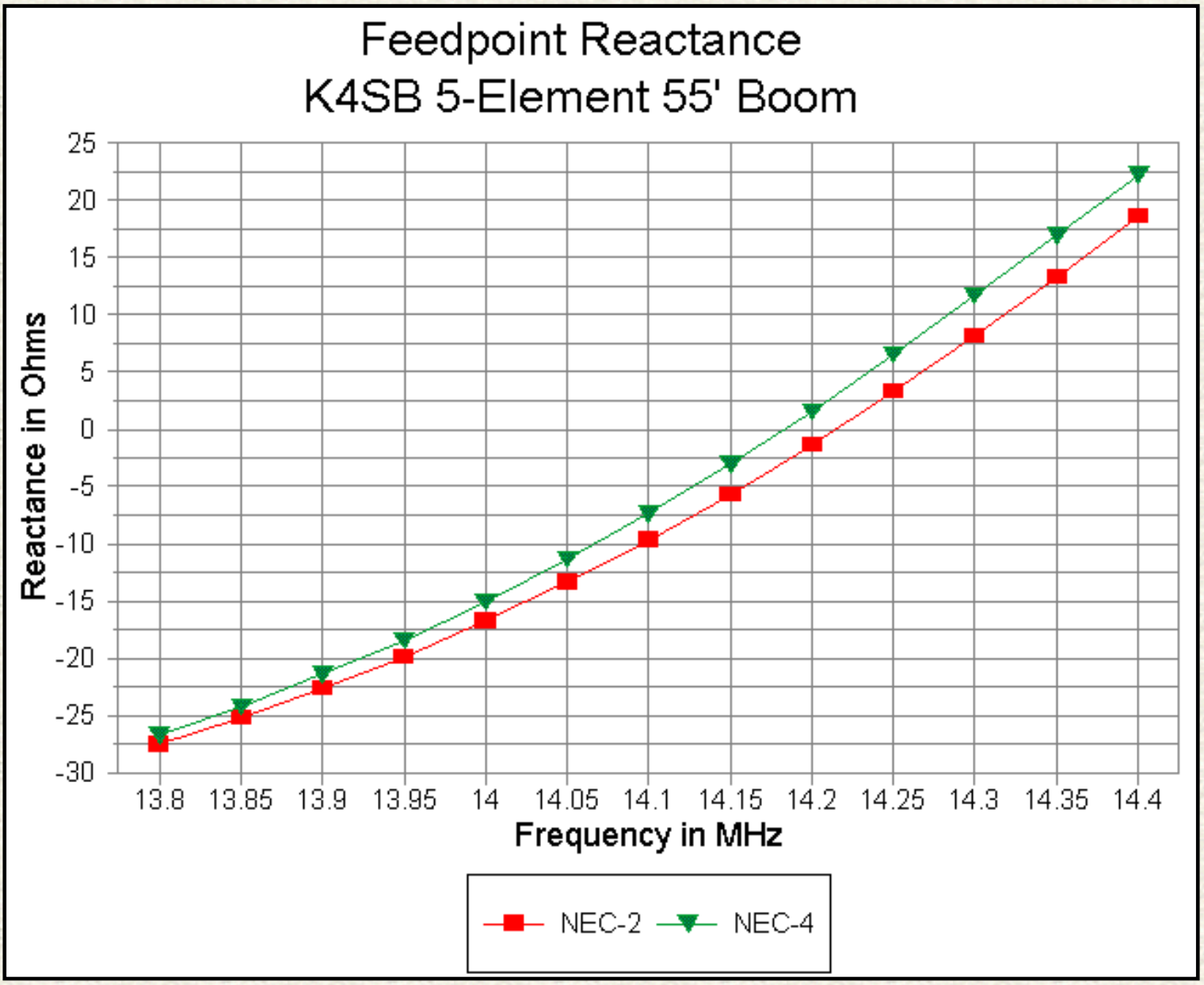
I received this model from K4SB (with whose permission it is used here) in simplified form, using monotapered elements on 10-meters. I frequency-scaled it to 20-meters and then assigned it an element taper schedule, readjusting the element lengths until it closely approximated the monotaper performance. The antenna shows the greatest gain and front-to-back differential across the band of all the models, and this exhibits the effects of using wide reflector spacing alone to raise the feedpoint impedance of the beam.

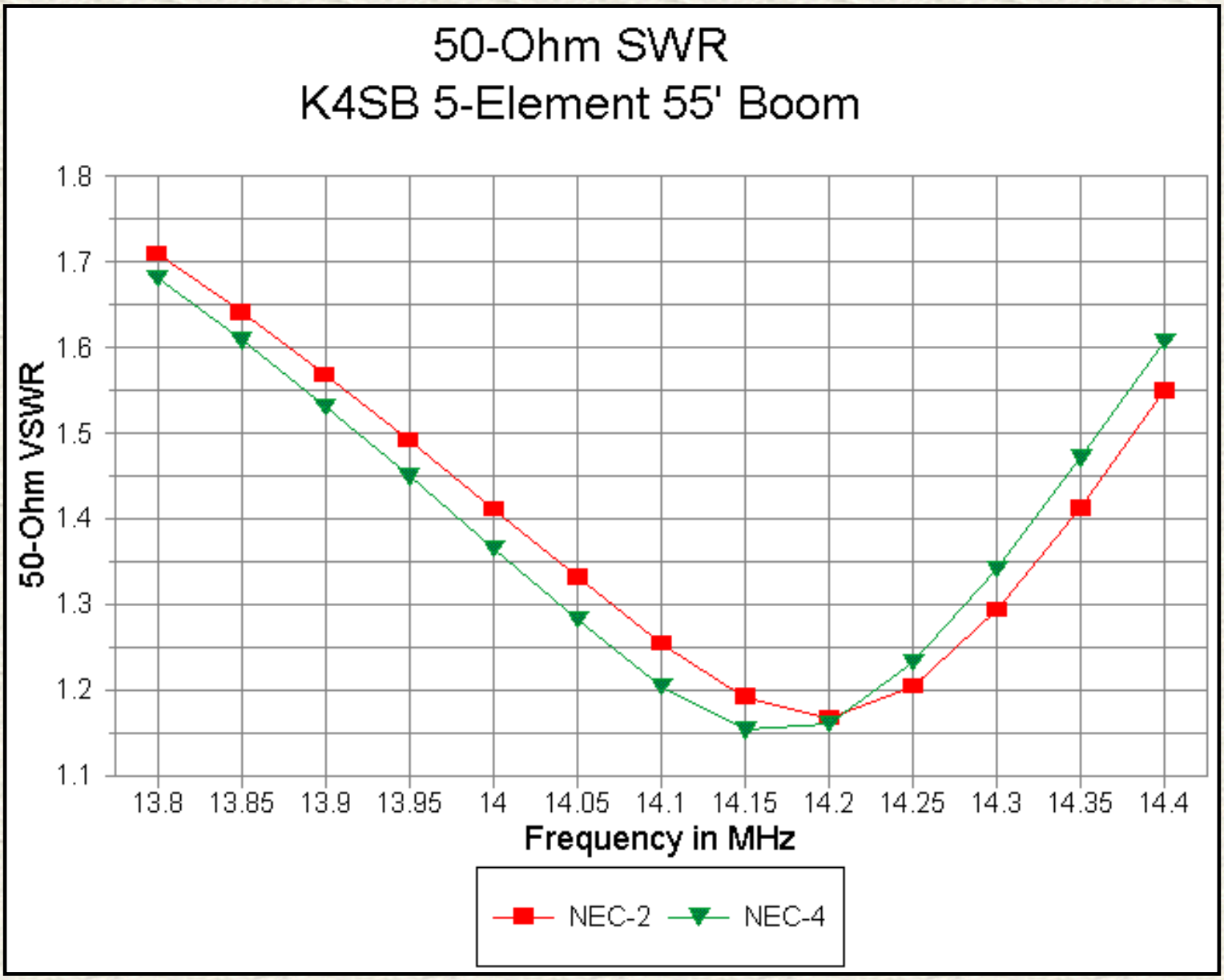
This beam is for me simply a modeling exercise and does not take into account any variations in building technique. It does not compensate for the effects of boom-to-element hardware. Considerable model tweaking would likely be required before construction.


Free Space Gain
K4SB 5-Element 55' Boom
Free Space Gain in dBi
10.8
10.6
10.4
10.2
10
9.8
9.6
9.4
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4


180-Degree Front-to-Back Ratio
K4SB 5-Element 55' Boom
180-Degree Front-to-Back Ratio in dB
60
50
40
30
20
10
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4

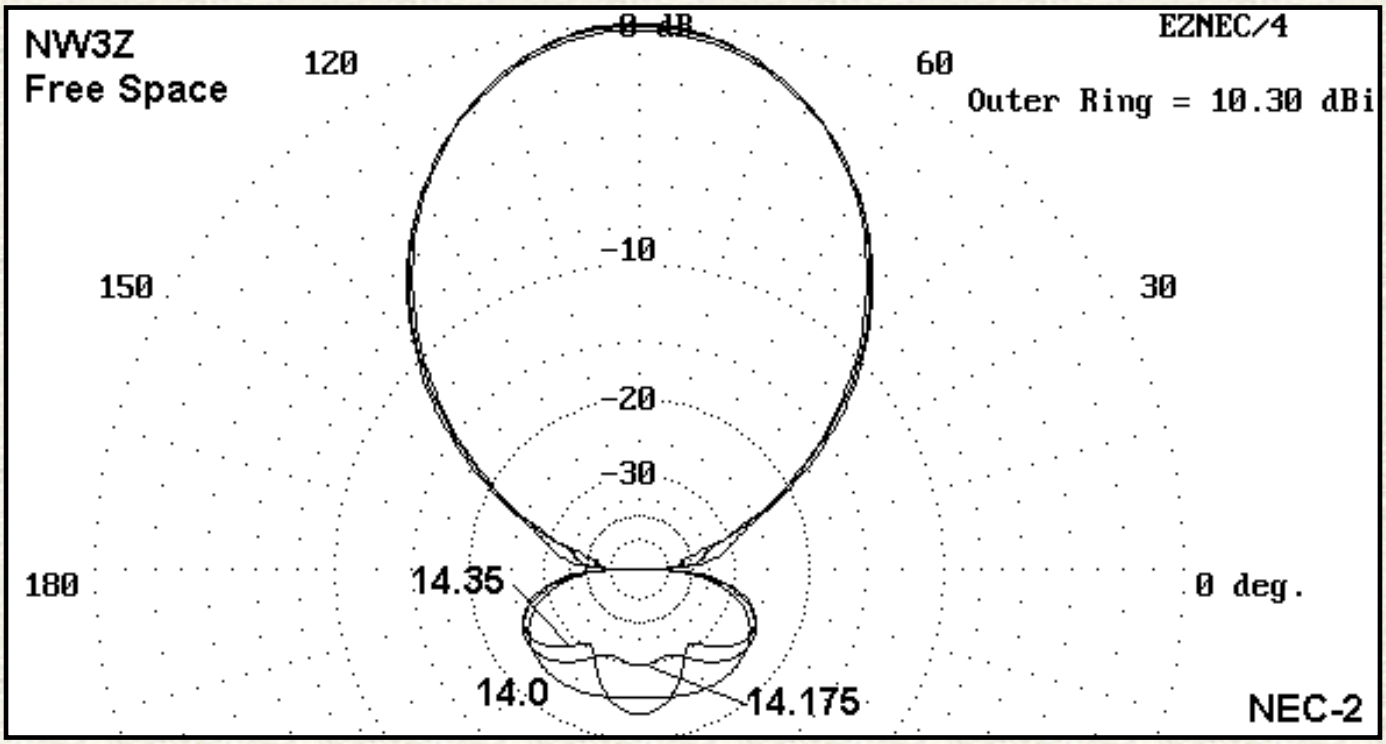

Feedpoint Resistance
K4SB 5-Element 55' Boom
Resistance in Ohms
55
50
45
40
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4


Feedpoint Reactance
K4SB 5-Element 55' Boom
Reactance in Ohms
25
20
15
10
5
0
-5
-10
-15
-20
-25
-30
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4



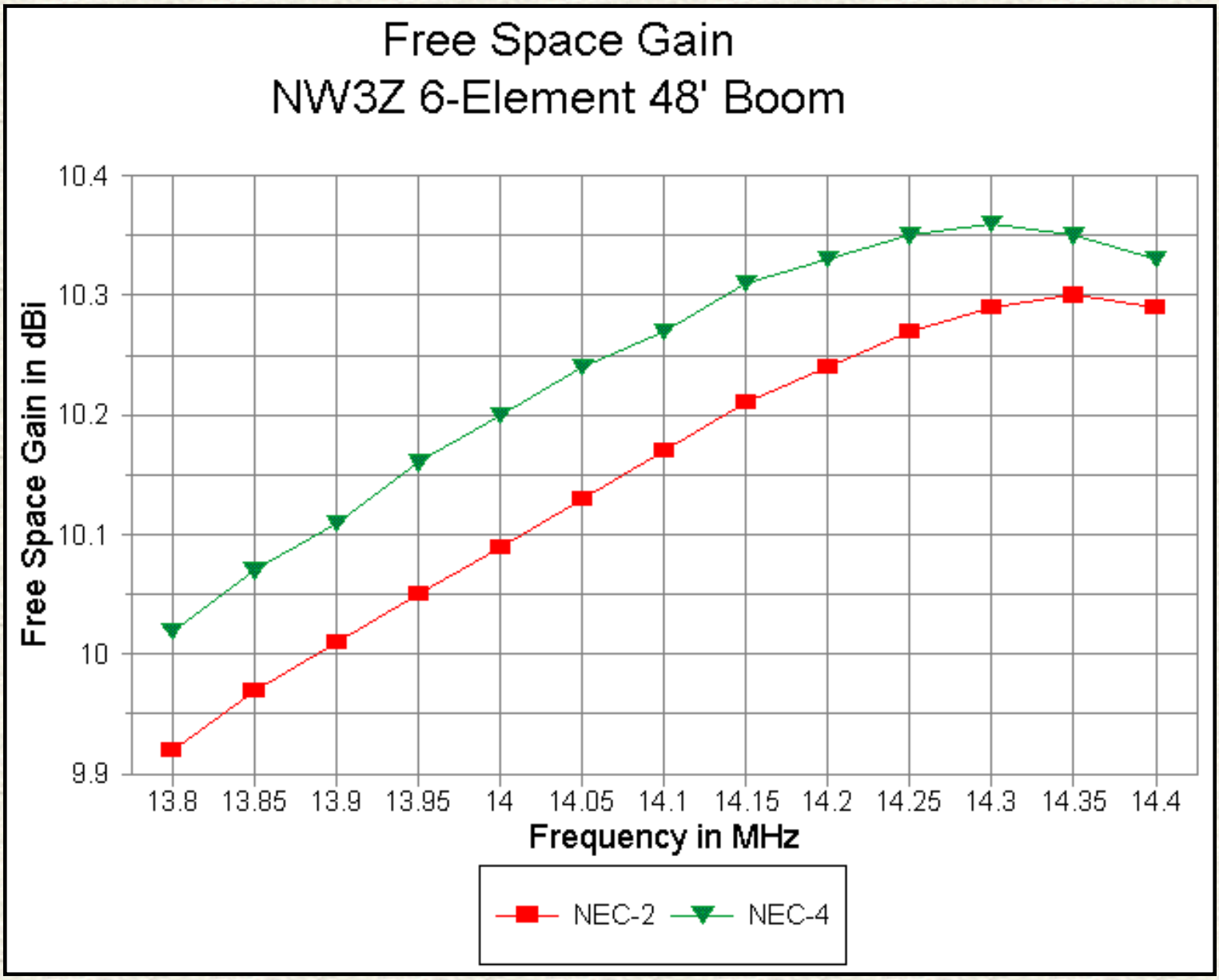
The SWR curve is quite good. There is a significant variation in reactance across the band, to go along with the variations in gain and front-to-back ratio, although the beam demonstrates close to the optimum of what may be achieved using refelctor spacing alone to determine a 50-Ohm feedpoint match.

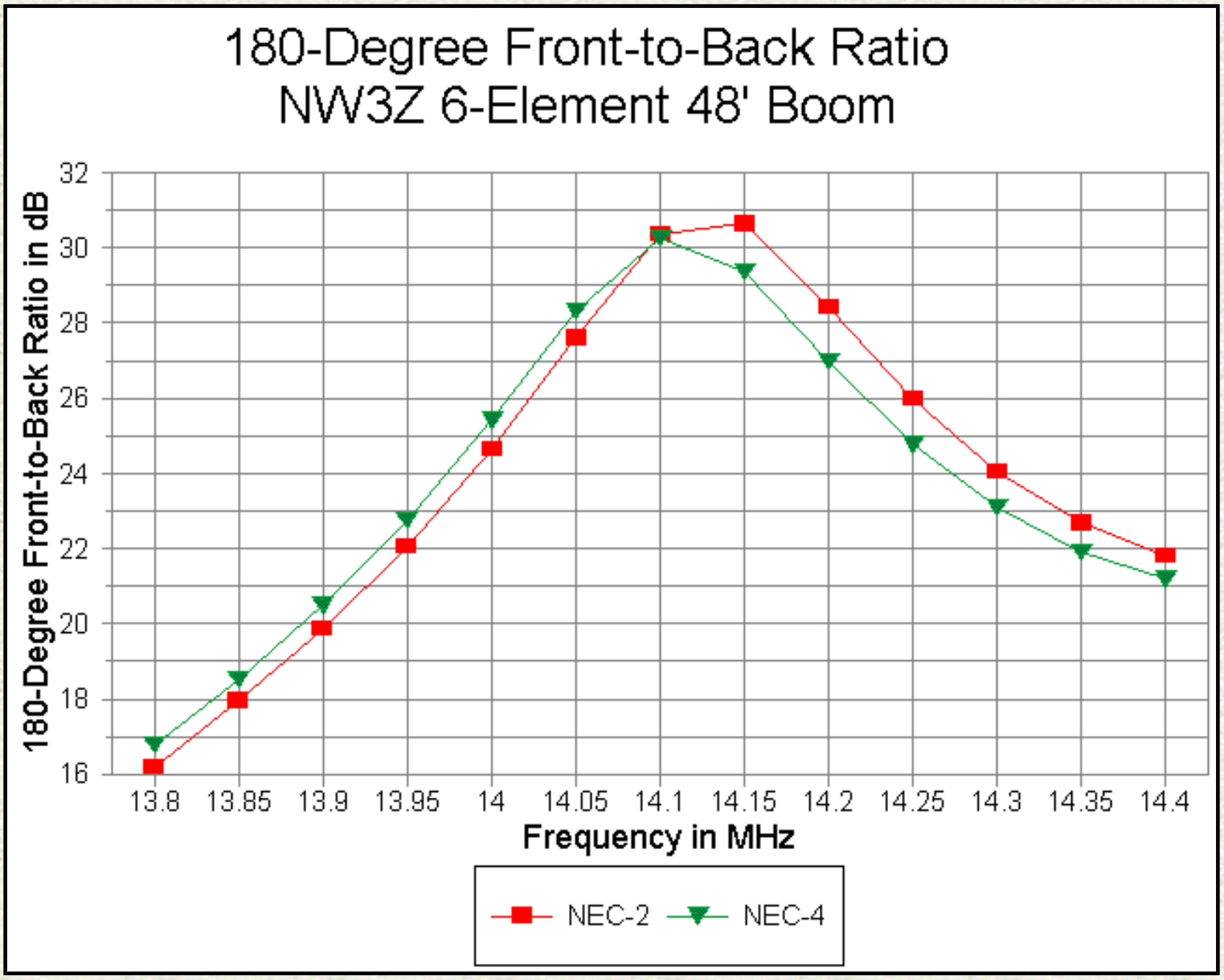
The NW3Z 6-element, 48' boom model

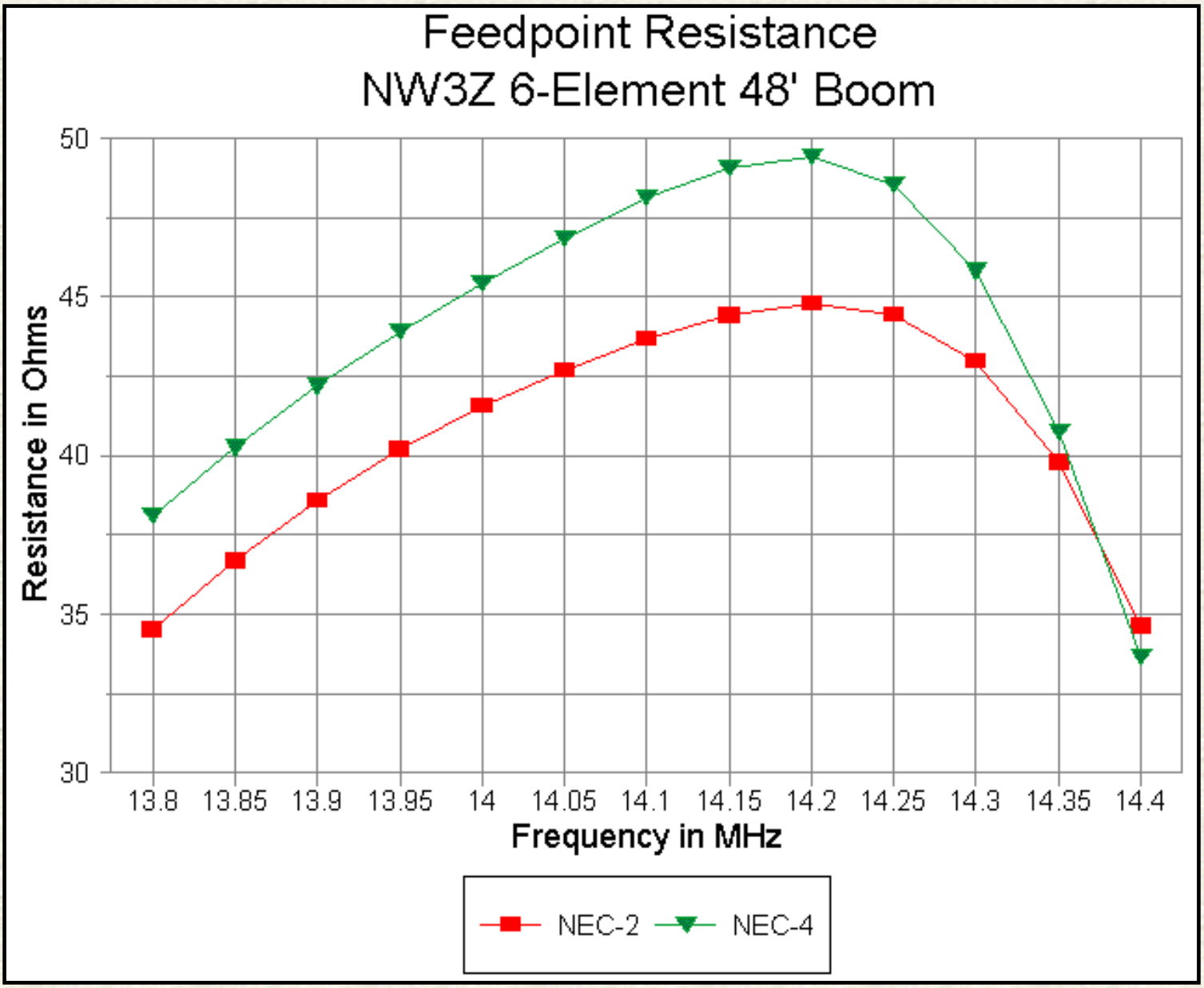


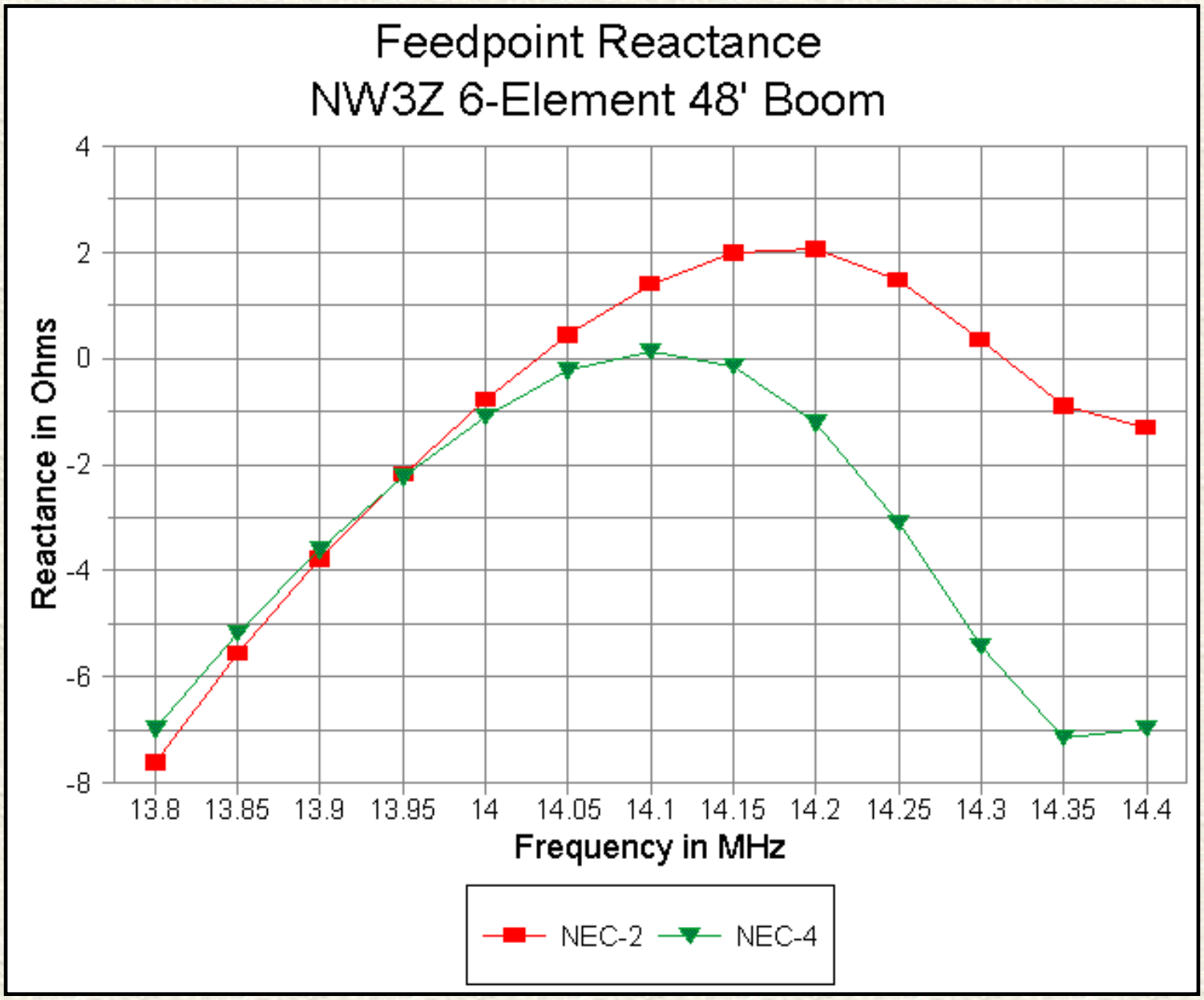
The NW3Z design adds a director close to the driven element. Along with the spacing of director D2 for impedance control and the spacing of the remaining directors for performance characteristics, this design provides a very tight near-50-Ohm feedpoint impedance with very little variation in other antenna performance characteristics across 20 meters. Indeed, everything remains stable for a little ways below the lower end of 20 meters.

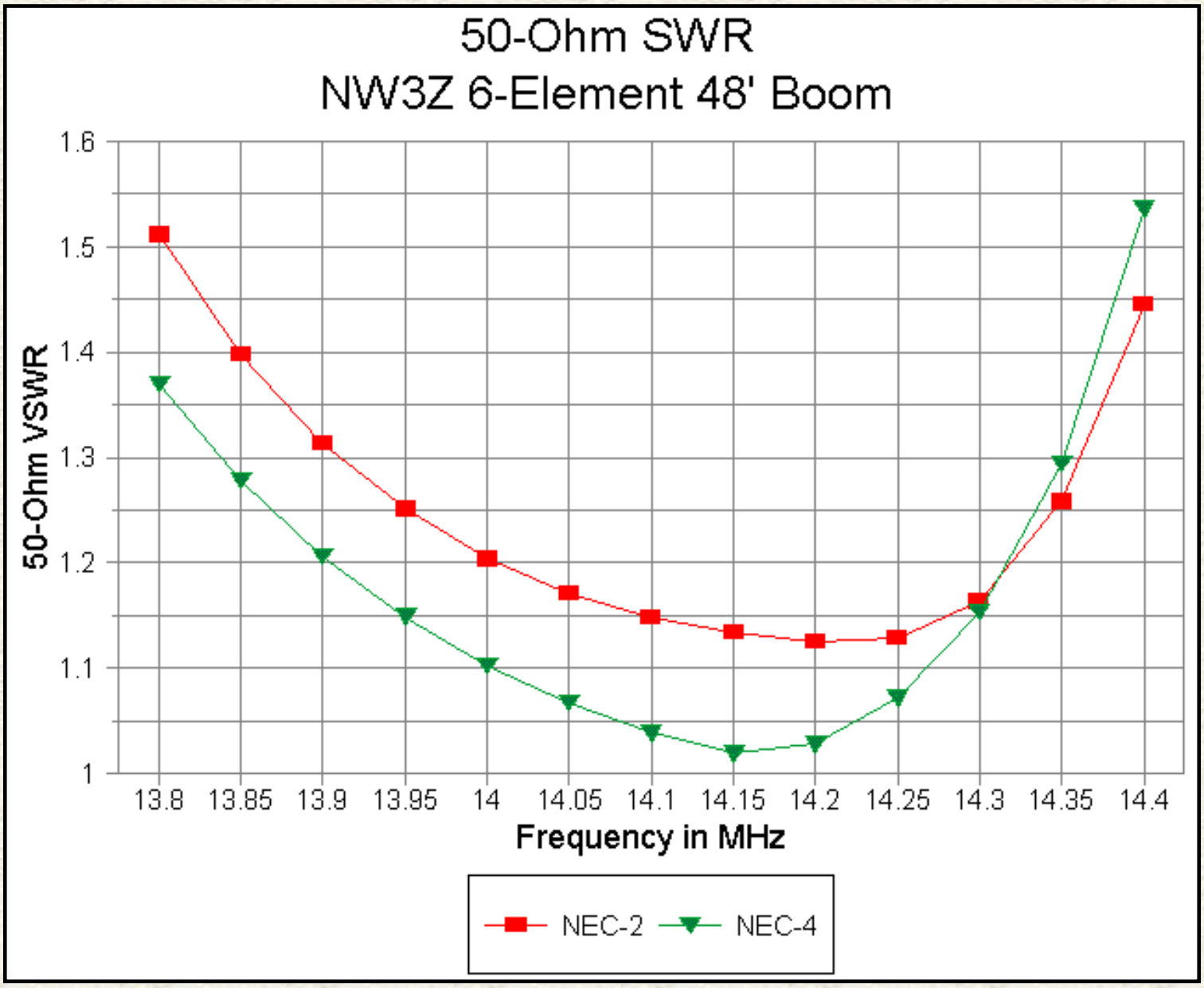
As with all other designs, NEC-4 tends to overestimate gain and move the performance characteristics lower in frequency relative to NEC-2 curves. Since this model uses a 96" constant diameter center section in each element, the effect is not too extreme, but still greater than for the W6NGZ design.


Free Space Gain
NW3Z 6-Element 48' Boom
Free Space Gain in dBi
10.4
10.3
10.2
10.1
10
9.9
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4


180-Degree Front-to-Back Ratio
NW3Z 6-Element 48' Boom
180-Degree Front-to-Back Ratio in dB
32
30
28
26
24
22
20
18
16
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4

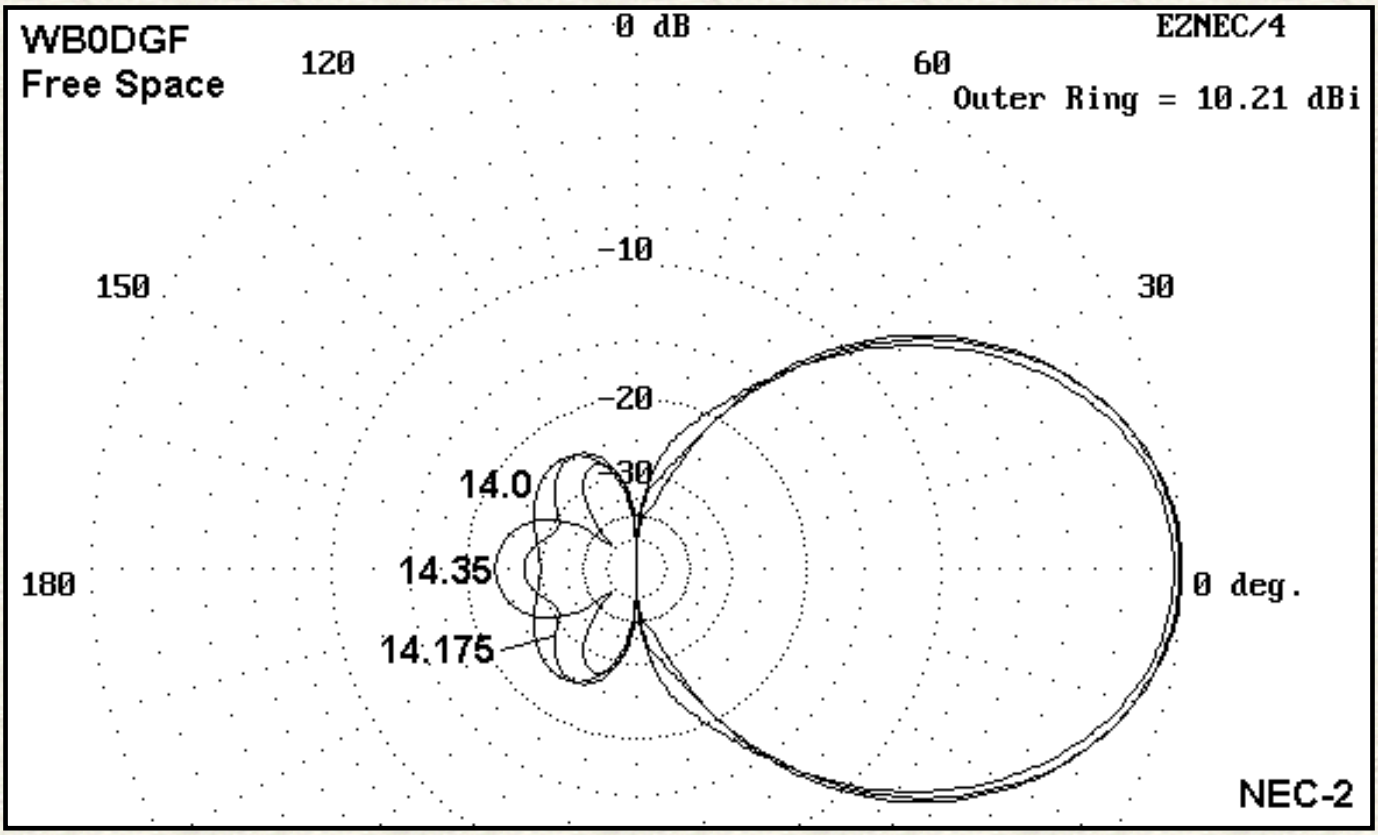

Feedpoint Resistance
NW3Z 6-Element 48' Boom
Resistance in Ohms
50
45
40
35
30
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4


Feedpoint Reactance
NW3Z 6-Element 48' Boom
Reactance in Ohms
4
2
0
-2
-4
-6
-8
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4



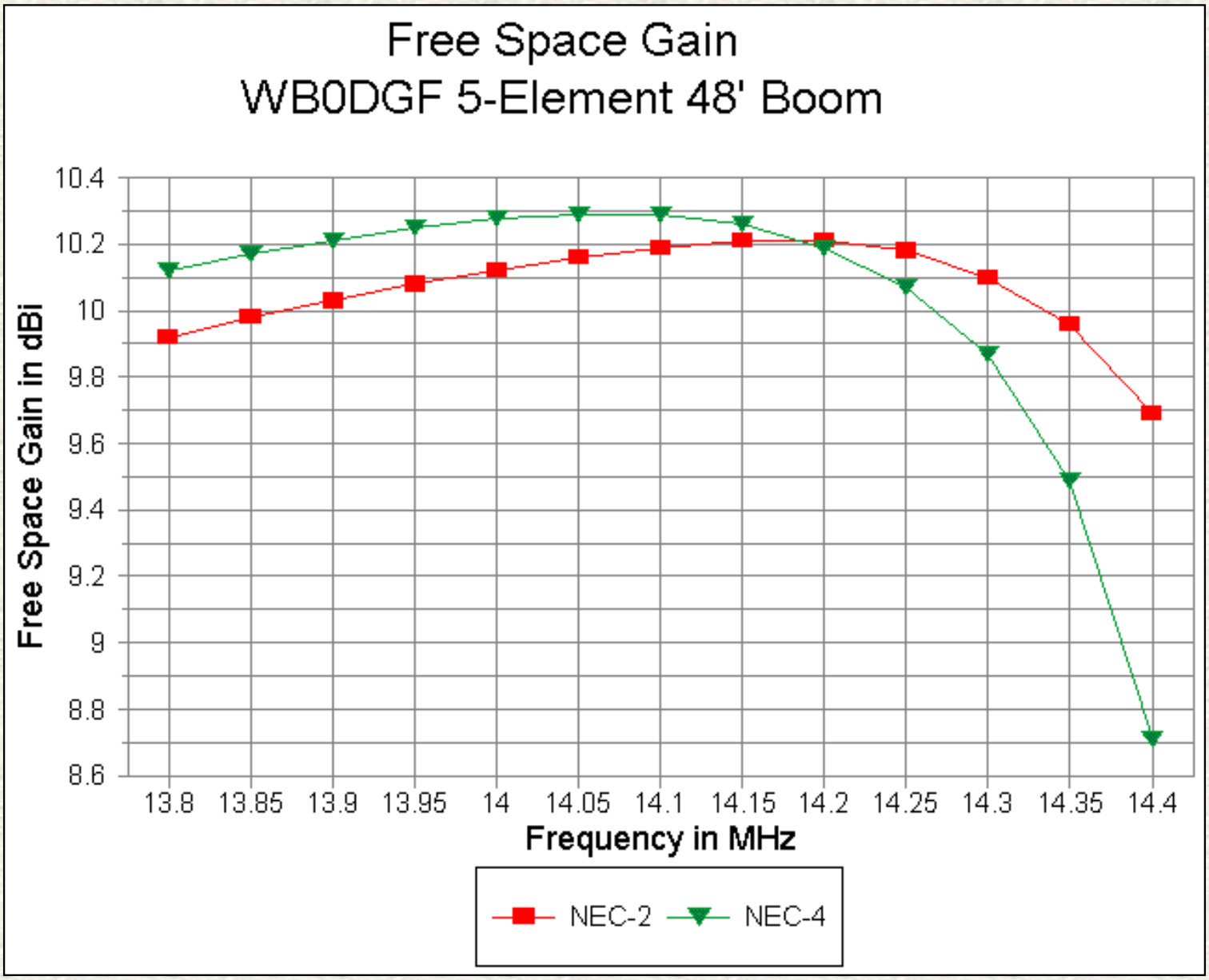
Beware of misreading the steep SWR curves: first read the axis values. Only then can you compare the numbers with those appearing in the other beam graphs.

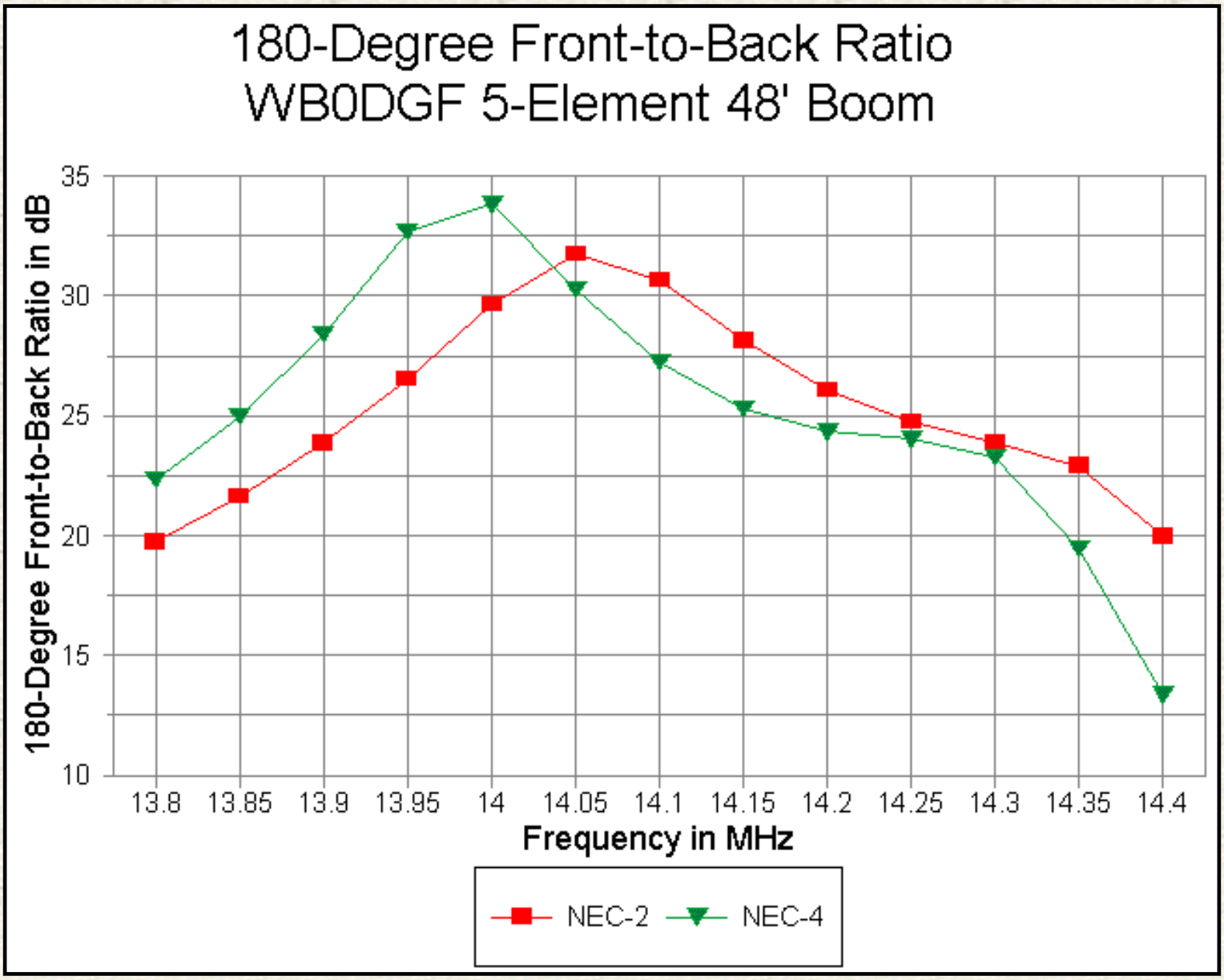
The WB0DGF 5-element, 48' boom model

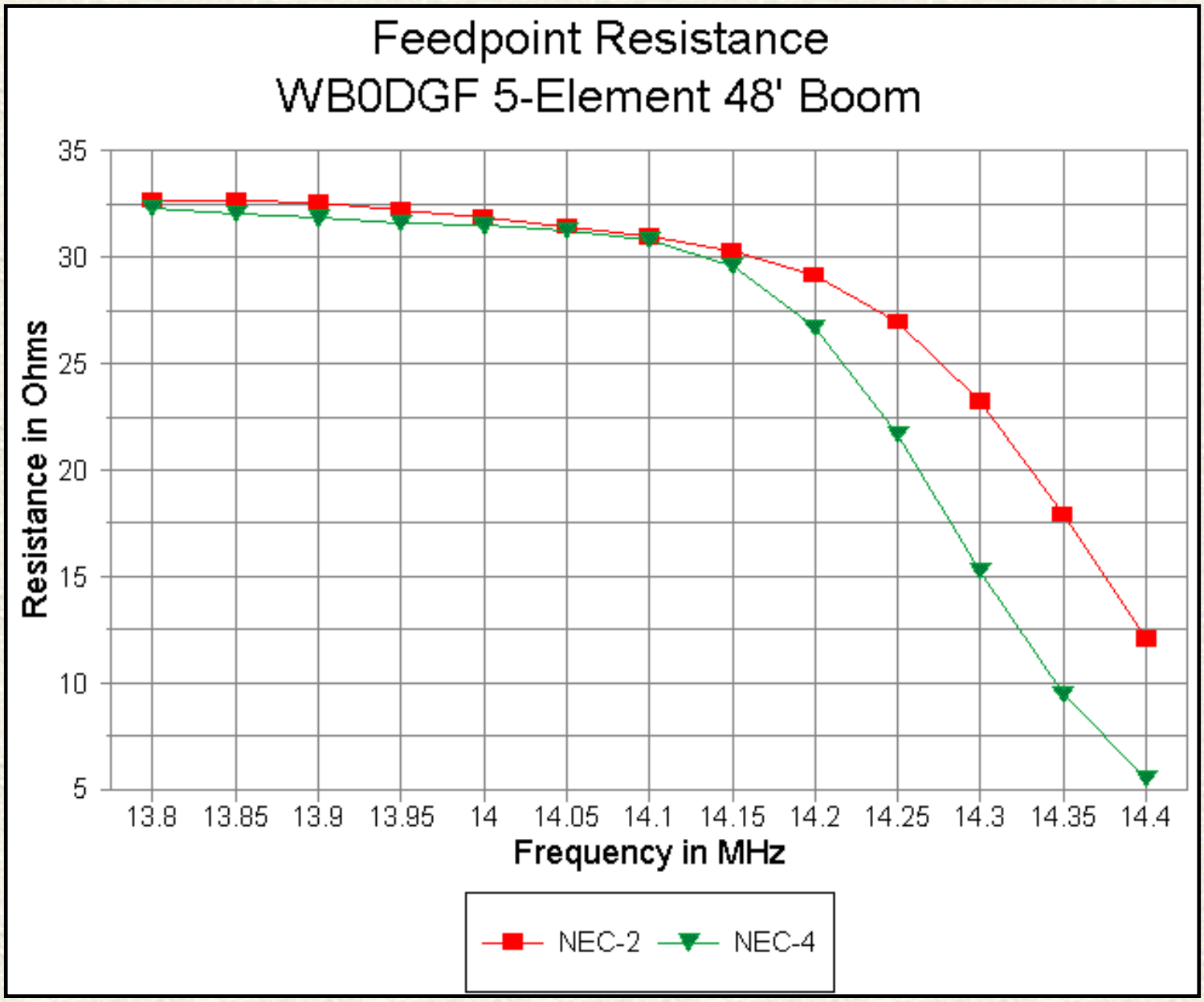


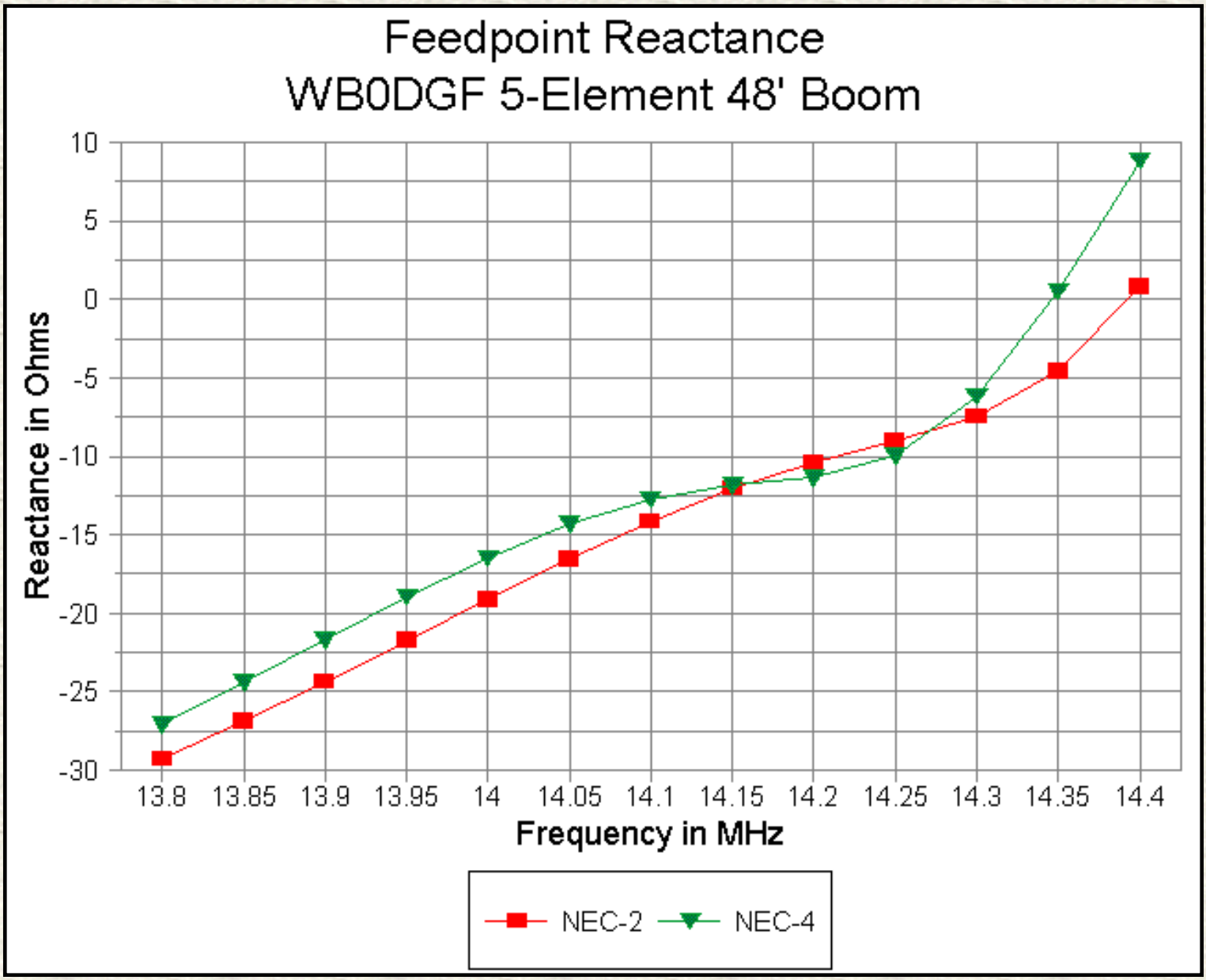
The WB0DGF design, provides an interesting contrast to both the K6STI and the NW3Z designs with respect to modeled properties. It contains center "bulge" sections that limit segment length to 8". However, these occur on only the parasitical elements. Moreover, the design uses a wider reflector spacing than some of the other designs. The result is a far less extreme overestimation of gain, but a significant shift in the NEC-4 curves lower in frequency than the NEC-2 curves. These phenomena are clear on all four of the graphs.

Since this antenna design was set up for potential use with a beta match system, no SWR curve is shown.


Free Space Gain
WB0DGF 5-Element 48' Boom
Free Space Gain in dBi
10.4
10.2
10
9.8
9.6
9.4
9.2
9
8.8
8.6
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4


180-Degree Front-to-Back Ratio
WB0DGF 5-Element 48' Boom
180-Degree Front-to-Back Ratio in dB
35
30
25
20
15
10
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4


Feedpoint Resistance
WB0DGF 5-Element 48' Boom
Resistance in Ohms
35
30
25
20
15
10
5
13.8
13.85
13.9
13.95
14
14.05
14.1
14.15
14.2
14.25
14.3
14.35
14.4
Frequency in MHz
NEC-2
NEC-4



Conclusions

Because of the open questions left along the way, no evaluative assessment is warranted by this exercise alone. However, two reminders are in order.

1. The beams do reveal something of the broader aspects of large Yagi behavior in the means we use to design specific feedpoint impedances while retaining the highest levels of gain and front-to-back performance. Study of the spacing sketch alone (made more precise by the EZNEC Pro file descriptions appended to this note) may give some insight into future Yagi design directions. Yagi design innovations are far from exhausted.

2. Modeling comparisons are not to be undertaken lightly or hastily. Analyzing the foundations of received models as well as those inherent in one's own creations is a necessary step if models are to be compared against some set of design or analysis goals. It would have been very easy to see two of the designs above as simply inadequate to cover the entirety of 20 meters by looking solely at NEC-4 curves, when what is more likely at work is a frequency offset. Add to this the limitations of each core used to model the structures unambiguously, and the simple conjunction of models would become a disservice to understanding the antennas involved.

Within these handicaps, the comparison of some of the best of extant long- boom 20-meter Yagi design provides an interesting and potentially productive exercise. It can assist us in understanding both antenna design and antenna modeling just a little bit better.

EZNEC Wire Tables for the 5 Yagis

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                    EZNEC/4  ver. 2.0

5L45' W6NGZ CQ 10-96 p 22                        07-19-1998     07:44:08


Frequency = 14.175  MHz.

Wire Loss: Aluminum -- Resistivity = 4E-08 ohm-m, Rel. Perm. = 1


              --------------- WIRES ---------------

Wire Conn. --- End 1 (x,y,z : in)  Conn. --- End 2 (x,y,z : in)   Dia(in) Segs

1          -215.60,  0.000,  0.000  W2E1 -156.00,  0.000,  0.000 6.25E-01   5
2     W1E2 -156.00,  0.000,  0.000  W3E1 -120.00,  0.000,  0.000 7.50E-01   3
3     W2E2 -120.00,  0.000,  0.000  W4E1 -72.000,  0.000,  0.000 8.75E-01   4
4     W3E2 -72.000,  0.000,  0.000  W5E1  72.000,  0.000,  0.000 1.00E+00  13
5     W4E2  72.000,  0.000,  0.000  W6E1 120.000,  0.000,  0.000 8.75E-01   4
6     W5E2 120.000,  0.000,  0.000  W7E1 156.000,  0.000,  0.000 7.50E-01   3
7     W6E2 156.000,  0.000,  0.000       215.605,  0.000,  0.000 6.25E-01   5
8          -205.95, 79.800,  0.000  W9E1 -156.00, 79.800,  0.000 6.25E-01   4
9     W8E2 -156.00, 79.800,  0.000 W10E1 -120.00, 79.800,  0.000 7.50E-01   3
10    W9E2 -120.00, 79.800,  0.000 W11E1 -72.000, 79.800,  0.000 8.75E-01   4
11   W10E2 -72.000, 79.800,  0.000 W12E1  72.000, 79.800,  0.000 1.00E+00  13
12   W11E2  72.000, 79.800,  0.000 W13E1 120.000, 79.800,  0.000 8.75E-01   4
13   W12E2 120.000, 79.800,  0.000 W14E1 156.000, 79.800,  0.000 7.50E-01   3
14   W13E2 156.000, 79.800,  0.000       205.950, 79.800,  0.000 6.25E-01   4
15         -198.21,155.160,  0.000 W16E1 -156.00,155.160,  0.000 6.25E-01   4
16   W15E2 -156.00,155.160,  0.000 W17E1 -120.00,155.160,  0.000 7.50E-01   3
17   W16E2 -120.00,155.160,  0.000 W18E1 -72.000,155.160,  0.000 8.75E-01   4
18   W17E2 -72.000,155.160,  0.000 W19E1  72.000,155.160,  0.000 1.00E+00  13
19   W18E2  72.000,155.160,  0.000 W20E1 120.000,155.160,  0.000 8.75E-01   4
20   W19E2 120.000,155.160,  0.000 W21E1 156.000,155.160,  0.000 7.50E-01   3
21   W20E2 156.000,155.160,  0.000       198.209,155.160,  0.000 6.25E-01   4
22         -196.55,337.920,  0.000 W23E1 -156.00,337.920,  0.000 6.25E-01   3
23   W22E2 -156.00,337.920,  0.000 W24E1 -120.00,337.920,  0.000 7.50E-01   3
24   W23E2 -120.00,337.920,  0.000 W25E1 -72.000,337.920,  0.000 8.75E-01   4
25   W24E2 -72.000,337.920,  0.000 W26E1  72.000,337.920,  0.000 1.00E+00  13
26   W25E2  72.000,337.920,  0.000 W27E1 120.000,337.920,  0.000 8.75E-01   4
27   W26E2 120.000,337.920,  0.000 W28E1 156.000,337.920,  0.000 7.50E-01   3
28   W27E2 156.000,337.920,  0.000       196.548,337.920,  0.000 6.25E-01   3
29         -189.90,530.400,  0.000 W30E1 -156.00,530.400,  0.000 6.25E-01   3
30   W29E2 -156.00,530.400,  0.000 W31E1 -120.00,530.400,  0.000 7.50E-01   3
```

31	W30E2	-120.00,530.400,	0.000	W32E1	-72.000,530.400,	0.000	8.75E-01	4
32	W31E2	-72.000,530.400,	0.000	W33E1	72.000,530.400,	0.000	1.00E+00	13
33	W32E2	72.000,530.400,	0.000	W34E1	120.000,530.400,	0.000	8.75E-01	3
34	W33E2	120.000,530.400,	0.000	W35E1	156.000,530.400,	0.000	7.50E-01	3
35	W34E2	156.000,530.400,	0.000		189.900,530.400,	0.000	6.25E-01	4

---------- SOURCES ----------

Source	Wire Seg.	Wire #/Pct From End 1 Actual	(Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	7	11 / 50.00	(11 / 50.00)	1.000	0.000	I

No loads specified

No transmission lines specified

Ground type is Free Space

EZNEC/4 ver. 2.0

5L48' K6STI YA 07-19-1998 07:44:45

Frequency = 14.175 MHz.

Wire Loss: Aluminum -- Resistivity = 4E-08 ohm-m, Rel. Perm. = 1

---------- WIRES ----------

Wire	Conn.	--- End 1 (x,y,z : in)			Conn.	--- End 2 (x,y,z : in)			Dia(in)	Segs
1		-216.50,	0.000,	0.000	W2E1	-138.00,	0.000,	0.000	5.00E-01	10
2	W1E2	-138.00,	0.000,	0.000	W3E1	-96.000,	0.000,	0.000	6.25E-01	6
3	W2E2	-96.000,	0.000,	0.000	W4E1	-48.000,	0.000,	0.000	7.50E-01	6
4	W3E2	-48.000,	0.000,	0.000	W5E1	-4.000,	0.000,	0.000	8.75E-01	5
5	W4E2	-4.000,	0.000,	0.000	W6E1	4.000,	0.000,	0.000	3.42E+00	1
6	W5E2	4.000,	0.000,	0.000	W7E1	48.000,	0.000,	0.000	8.75E-01	5
7	W6E2	48.000,	0.000,	0.000	W8E1	96.000,	0.000,	0.000	7.50E-01	6
8	W7E2	96.000,	0.000,	0.000	W9E1	138.000,	0.000,	0.000	6.25E-01	6
9	W8E2	138.000,	0.000,	0.000		216.500,	0.000,	0.000	5.00E-01	10
10		-203.50,	72.000,	0.000	W11E1	-138.00,	72.000,	0.000	5.00E-01	8
11	W10E2	-138.00,	72.000,	0.000	W12E1	-96.000,	72.000,	0.000	6.25E-01	6
12	W11E2	-96.000,	72.000,	0.000	W13E1	-48.000,	72.000,	0.000	7.50E-01	6
13	W12E2	-48.000,	72.000,	0.000	W14E1	-4.000,	72.000,	0.000	8.75E-01	5
14	W13E2	-4.000,	72.000,	0.000	W15E1	4.000,	72.000,	0.000	3.42E+00	1
15	W14E2	4.000,	72.000,	0.000	W16E1	48.000,	72.000,	0.000	8.75E-01	5
16	W15E2	48.000,	72.000,	0.000	W17E1	96.000,	72.000,	0.000	7.50E-01	6
17	W16E2	96.000,	72.000,	0.000	W18E1	138.000,	72.000,	0.000	6.25E-01	6

18	W17E2	138.000,	72.000,	0.000		203.500,	72.000,	0.000	5.00E-01	8
19		-201.75,	160.000,	0.000	W20E1	-138.00,	160.000,	0.000	5.00E-01	8
20	W19E2	-138.00,	160.000,	0.000	W21E1	-96.000,	160.000,	0.000	6.25E-01	6
21	W20E2	-96.000,	160.000,	0.000	W22E1	-48.000,	160.000,	0.000	7.50E-01	6
22	W21E2	-48.000,	160.000,	0.000	W23E1	-4.000,	160.000,	0.000	8.75E-01	5
23	W22E2	-4.000,	160.000,	0.000	W24E1	4.000,	160.000,	0.000	3.42E+00	1
24	W23E2	4.000,	160.000,	0.000	W25E1	48.000,	160.000,	0.000	8.75E-01	5
25	W24E2	48.000,	160.000,	0.000	W26E1	96.000,	160.000,	0.000	7.50E-01	6
26	W25E2	96.000,	160.000,	0.000	W27E1	138.000,	160.000,	0.000	6.25E-01	6
27	W26E2	138.000,	160.000,	0.000		201.750,	160.000,	0.000	5.00E-01	8
28		-198.88,	359.000,	0.000	W29E1	-138.00,	359.000,	0.000	5.00E-01	8
29	W28E2	-138.00,	359.000,	0.000	W30E1	-96.000,	359.000,	0.000	6.25E-01	5
30	W29E2	-96.000,	359.000,	0.000	W31E1	-48.000,	359.000,	0.000	7.50E-01	6
31	W30E2	-48.000,	359.000,	0.000	W32E1	-4.000,	359.000,	0.000	8.75E-01	6
32	W31E2	-4.000,	359.000,	0.000	W33E1	4.000,	359.000,	0.000	3.42E+00	1
33	W32E2	4.000,	359.000,	0.000	W34E1	48.000,	359.000,	0.000	8.75E-01	6
34	W33E2	48.000,	359.000,	0.000	W35E1	96.000,	359.000,	0.000	7.50E-01	6
35	W34E2	96.000,	359.000,	0.000	W36E1	138.000,	359.000,	0.000	6.25E-01	5
36	W35E2	138.000,	359.000,	0.000		198.875,	359.000,	0.000	5.00E-01	8
37		-191.62,	570.000,	0.000	W38E1	-138.00,	570.000,	0.000	5.00E-01	7
38	W37E2	-138.00,	570.000,	0.000	W39E1	-96.000,	570.000,	0.000	6.25E-01	5
39	W38E2	-96.000,	570.000,	0.000	W40E1	-48.000,	570.000,	0.000	7.50E-01	6
40	W39E2	-48.000,	570.000,	0.000	W41E1	-4.000,	570.000,	0.000	8.75E-01	6
41	W40E2	-4.000,	570.000,	0.000	W42E1	4.000,	570.000,	0.000	3.42E+00	1
42	W41E2	4.000,	570.000,	0.000	W43E1	48.000,	570.000,	0.000	8.75E-01	6
43	W42E2	48.000,	570.000,	0.000	W44E1	96.000,	570.000,	0.000	7.50E-01	6
44	W43E2	96.000,	570.000,	0.000	W45E1	138.000,	570.000,	0.000	6.25E-01	5
45	W44E2	138.000,	570.000,	0.000		191.625,	570.000,	0.000	5.00E-01	7

---------- SOURCES ----------

Source	Wire Seg.	Wire #/Pct From End 1 Actual	(Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	1	14 / 50.00	(14 / 50.00)	1.000	0.000	V

No loads specified

No transmission lines specified

Ground type is Free Space

EZNEC/4 ver. 2.0

5L48' W3LPL WWW 07-19-1998 07:45:37

Frequency = 14.175 MHz.

Wire Loss: Aluminum -- Resistivity = 4E-08 ohm-m, Rel. Perm. = 1

--------------- WIRES ---------------

Wire	Conn.	--- End 1 (x,y,z : in)			Conn.	--- End 2 (x,y,z : in)			Dia(in)	Segs
1		-217.00,	0.000,	0.000	W2E1	-150.00,	0.000,	0.000	7.50E-01	8
2	W1E2	-150.00,	0.000,	0.000	W3E1	-117.00,	0.000,	0.000	8.75E-01	4
3	W2E2	-117.00,	0.000,	0.000	W4E1	-84.000,	0.000,	0.000	1.00E+00	4
4	W3E2	-84.000,	0.000,	0.000	W5E1	-33.750,	0.000,	0.000	1.12E+00	6
5	W4E2	-33.750,	0.000,	0.000	W6E1	33.750,	0.000,	0.000	1.25E+00	9
6	W5E2	33.750,	0.000,	0.000	W7E1	84.000,	0.000,	0.000	1.12E+00	6
7	W6E2	84.000,	0.000,	0.000	W8E1	117.000,	0.000,	0.000	1.00E+00	4
8	W7E2	117.000,	0.000,	0.000	W9E1	150.000,	0.000,	0.000	8.75E-01	4
9	W8E2	150.000,	0.000,	0.000		217.000,	0.000,	0.000	7.50E-01	8
10		-212.00,	86.000,	0.000	W11E1	-150.00,	86.000,	0.000	7.50E-01	8
11	W10E2	-150.00,	86.000,	0.000	W12E1	-117.00,	86.000,	0.000	8.75E-01	4
12	W11E2	-117.00,	86.000,	0.000	W13E1	-84.000,	86.000,	0.000	1.00E+00	4
13	W12E2	-84.000,	86.000,	0.000	W14E1	-33.750,	86.000,	0.000	1.12E+00	6
14	W13E2	-33.750,	86.000,	0.000	W15E1	33.750,	86.000,	0.000	1.25E+00	9
15	W14E2	33.750,	86.000,	0.000	W16E1	84.000,	86.000,	0.000	1.12E+00	6
16	W15E2	84.000,	86.000,	0.000	W17E1	117.000,	86.000,	0.000	1.00E+00	4
17	W16E2	117.000,	86.000,	0.000	W18E1	150.000,	86.000,	0.000	8.75E-01	4
18	W17E2	150.000,	86.000,	0.000		212.000,	86.000,	0.000	7.50E-01	8
19		-199.38,	171.000,	0.000	W20E1	-135.88,	171.000,	0.000	7.50E-01	8
20	W19E2	-135.88,	171.000,	0.000	W21E1	-102.88,	171.000,	0.000	8.75E-01	4
21	W20E2	-102.88,	171.000,	0.000	W22E1	-69.875,	171.000,	0.000	1.00E+00	4
22	W21E2	-69.875,	171.000,	0.000	W23E1	-22.000,	171.000,	0.000	1.12E+00	6
23	W22E2	-22.000,	171.000,	0.000	W24E1	22.000,	171.000,	0.000	1.25E+00	5
24	W23E2	22.000,	171.000,	0.000	W25E1	69.875,	171.000,	0.000	1.12E+00	6
25	W24E2	69.875,	171.000,	0.000	W26E1	102.875,	171.000,	0.000	1.00E+00	4
26	W25E2	102.875,	171.000,	0.000	W27E1	135.875,	171.000,	0.000	8.75E-01	4
27	W26E2	135.875,	171.000,	0.000		199.380,	171.000,	0.000	7.50E-01	8
28		-196.25,	349.000,	0.000	W29E1	-119.25,	349.000,	0.000	7.50E-01	10
29	W28E2	-119.25,	349.000,	0.000	W30E1	-86.250,	349.000,	0.000	8.75E-01	4
30	W29E2	-86.250,	349.000,	0.000	W31E1	-53.250,	349.000,	0.000	1.00E+00	4
31	W30E2	-53.250,	349.000,	0.000	W32E1	-22.000,	349.000,	0.000	1.12E+00	4
32	W31E2	-22.000,	349.000,	0.000	W33E1	22.000,	349.000,	0.000	1.25E+00	5
33	W32E2	22.000,	349.000,	0.000	W34E1	53.250,	349.000,	0.000	1.12E+00	4
34	W33E2	53.250,	349.000,	0.000	W35E1	86.250,	349.000,	0.000	1.00E+00	4
35	W34E2	86.250,	349.000,	0.000	W36E1	119.250,	349.000,	0.000	8.75E-01	4
36	W35E2	119.250,	349.000,	0.000		196.250,	349.000,	0.000	7.50E-01	10
37		-183.75,	570.000,	0.000	W38E1	-131.75,	570.000,	0.000	7.50E-01	7
38	W37E2	-131.75,	570.000,	0.000	W39E1	-98.750,	570.000,	0.000	8.75E-01	4
39	W38E2	-98.750,	570.000,	0.000	W40E1	-65.750,	570.000,	0.000	1.00E+00	4
40	W39E2	-65.750,	570.000,	0.000	W41E1	-22.000,	570.000,	0.000	1.12E+00	5
41	W40E2	-22.000,	570.000,	0.000	W42E1	22.000,	570.000,	0.000	1.25E+00	5
42	W41E2	22.000,	570.000,	0.000	W43E1	65.750,	570.000,	0.000	1.12E+00	5
43	W42E2	65.750,	570.000,	0.000	W44E1	98.750,	570.000,	0.000	1.00E+00	4

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44   W43E2  98.750,570.000,  0.000 W45E1 131.750,570.000,  0.000 8.75E-01   4
45   W44E2 131.750,570.000,  0.000       183.750,570.000,  0.000 7.50E-01   7

             -------------- SOURCES --------------

Source    Wire      Wire #/Pct From End 1    Ampl.(V, A)  Phase(Deg.)  Type
          Seg.     Actual      (Specified)

1           5    14 / 50.00   ( 14 / 50.00)      1.000       0.000       V

No loads specified

No transmission lines specified

Ground type is Free Space
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                      EZNEC/4  ver. 2.0

K4SB 5-element, 55' boom                        07-19-1998     07:46:17


Frequency = 14.175  MHz.

Wire Loss: Aluminum -- Resistivity = 4E-08 ohm-m, Rel. Perm. = 1


              --------------- WIRES ---------------

Wire Conn. --- End 1 (x,y,z : in)  Conn. --- End 2 (x,y,z : in)   Dia(in) Segs

1          -217.00,  0.000,  0.000  W2E1 -152.00,  0.000,  0.000 5.00E-01   7
2     W1E2 -152.00,  0.000,  0.000  W3E1 -116.00,  0.000,  0.000 6.25E-01   4
3     W2E2 -116.00,  0.000,  0.000  W4E1 -72.000,  0.000,  0.000 7.50E-01   4
4     W3E2 -72.000,  0.000,  0.000  W5E1 -48.000,  0.000,  0.000 8.75E-01   2
5     W4E2 -48.000,  0.000,  0.000  W6E1  48.000,  0.000,  0.000 1.00E+00   9
6     W5E2  48.000,  0.000,  0.000  W7E1  72.000,  0.000,  0.000 8.75E-01   2
7     W6E2  72.000,  0.000,  0.000  W8E1 116.000,  0.000,  0.000 7.50E-01   4
8     W7E2 116.000,  0.000,  0.000  W9E1 152.000,  0.000,  0.000 6.25E-01   4
9     W8E2 152.000,  0.000,  0.000       217.000,  0.000,  0.000 5.00E-01   7
10         -207.50,182.504,  0.000 W11E1 -152.00,182.504,  0.000 5.00E-01   6
11   W10E2 -152.00,182.504,  0.000 W12E1 -116.00,182.504,  0.000 6.25E-01   4
12   W11E2 -116.00,182.504,  0.000 W13E1 -72.000,182.504,  0.000 7.50E-01   4
13   W12E2 -72.000,182.504,  0.000 W14E1 -48.000,182.504,  0.000 8.75E-01   2
14   W13E2 -48.000,182.504,  0.000 W15E1  48.000,182.504,  0.000 1.00E+00   9
15   W14E2  48.000,182.504,  0.000 W16E1  72.000,182.504,  0.000 8.75E-01   2
16   W15E2  72.000,182.504,  0.000 W17E1 116.000,182.504,  0.000 7.50E-01   4
17   W16E2 116.000,182.504,  0.000 W18E1 152.000,182.504,  0.000 6.25E-01   4
18   W17E2 152.000,182.504,  0.000       207.500,182.504,  0.000 5.00E-01   6
19         -195.53,269.544,  0.000 W20E1 -152.00,269.544,  0.000 5.00E-01   4
20   W19E2 -152.00,269.544,  0.000 W21E1 -116.00,269.544,  0.000 6.25E-01   4
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21   W20E2 -116.00,269.544,  0.000 W22E1 -72.000,269.544,  0.000 7.50E-01   4
22   W21E2 -72.000,269.544,  0.000 W23E1 -48.000,269.544,  0.000 8.75E-01   2
23   W22E2 -48.000,269.544,  0.000 W24E1  48.000,269.544,  0.000 1.00E+00   9
24   W23E2  48.000,269.544,  0.000 W25E1  72.000,269.544,  0.000 8.75E-01   2
25   W24E2  72.000,269.544,  0.000 W26E1 116.000,269.544,  0.000 7.50E-01   4
26   W25E2 116.000,269.544,  0.000 W27E1 152.000,269.544,  0.000 6.25E-01   4
27   W26E2 152.000,269.544,  0.000       195.530,269.544,  0.000 5.00E-01   4
28         -192.48,448.494,  0.000 W29E1 -152.00,448.494,  0.000 5.00E-01   4
29   W28E2 -152.00,448.494,  0.000 W30E1 -116.00,448.494,  0.000 6.25E-01   4
30   W29E2 -116.00,448.494,  0.000 W31E1 -72.000,448.494,  0.000 7.50E-01   4
31   W30E2 -72.000,448.494,  0.000 W32E1 -48.000,448.494,  0.000 8.75E-01   2
32   W31E2 -48.000,448.494,  0.000 W33E1  48.000,448.494,  0.000 1.00E+00   9
33   W32E2  48.000,448.494,  0.000 W34E1  72.000,448.494,  0.000 8.75E-01   2
34   W33E2  72.000,448.494,  0.000 W35E1 116.000,448.494,  0.000 7.50E-01   4
35   W34E2 116.000,448.494,  0.000 W36E1 152.000,448.494,  0.000 6.25E-01   4
36   W35E2 152.000,448.494,  0.000       192.480,448.494,  0.000 5.00E-01   4
37         -189.42,656.474,  0.000 W38E1 -152.00,656.474,  0.000 5.00E-01   4
38   W37E2 -152.00,656.474,  0.000 W39E1 -116.00,656.474,  0.000 6.25E-01   4
39   W38E2 -116.00,656.474,  0.000 W40E1 -72.000,656.474,  0.000 7.50E-01   4
40   W39E2 -72.000,656.474,  0.000 W41E1 -48.000,656.474,  0.000 8.75E-01   2
41   W40E2 -48.000,656.474,  0.000 W42E1  48.000,656.474,  0.000 1.00E+00   9
42   W41E2  48.000,656.474,  0.000 W43E1  72.000,656.474,  0.000 8.75E-01   2
43   W42E2  72.000,656.474,  0.000 W44E1 116.000,656.474,  0.000 7.50E-01   4
44   W43E2 116.000,656.474,  0.000 W45E1 152.000,656.474,  0.000 6.25E-01   4
45   W44E2 152.000,656.474,  0.000       189.420,656.474,  0.000 5.00E-01   4

             -------------- SOURCES --------------

Source    Wire      Wire #/Pct From End 1    Ampl.(V, A)  Phase(Deg.)  Type
          Seg.     Actual      (Specified)

1           5    14 / 50.00   ( 14 / 50.00)      1.000        0.000       V

No loads specified

No transmission lines specified

Ground type is Free Space
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                     EZNEC/4  ver. 2.0

NW3Z 6 el, 48' boom WWW                      07-19-1998     07:47:18


Frequency = 14.175  MHz.

Wire Loss: Aluminum -- Resistivity = 4E-08 ohm-m, Rel. Perm. = 1


             --------------- WIRES ---------------
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Wire	Conn.	--- End 1 (x,y,z : in)			Conn.	--- End 2 (x,y,z : in)			Dia(in)	Segs
1		-217.73,	0.000,	0.000	W2E1	-152.00,	0.000,	0.000	5.00E-01	7
2	W1E2	-152.00,	0.000,	0.000	W3E1	-116.00,	0.000,	0.000	6.25E-01	4
3	W2E2	-116.00,	0.000,	0.000	W4E1	-72.000,	0.000,	0.000	7.50E-01	4
4	W3E2	-72.000,	0.000,	0.000	W5E1	-48.000,	0.000,	0.000	8.75E-01	2
5	W4E2	-48.000,	0.000,	0.000	W6E1	48.000,	0.000,	0.000	1.00E+00	9
6	W5E2	48.000,	0.000,	0.000	W7E1	72.000,	0.000,	0.000	8.75E-01	2
7	W6E2	72.000,	0.000,	0.000	W8E1	116.000,	0.000,	0.000	7.50E-01	4
8	W7E2	116.000,	0.000,	0.000	W9E1	152.000,	0.000,	0.000	6.25E-01	4
9	W8E2	152.000,	0.000,	0.000		217.730,	0.000,	0.000	5.00E-01	7
10		-210.70,	90.000,	0.000	W11E1	-152.00,	90.000,	0.000	5.00E-01	6
11	W10E2	-152.00,	90.000,	0.000	W12E1	-116.00,	90.000,	0.000	6.25E-01	4
12	W11E2	-116.00,	90.000,	0.000	W13E1	-72.000,	90.000,	0.000	7.50E-01	4
13	W12E2	-72.000,	90.000,	0.000	W14E1	-48.000,	90.000,	0.000	8.75E-01	2
14	W13E2	-48.000,	90.000,	0.000	W15E1	48.000,	90.000,	0.000	1.00E+00	9
15	W14E2	48.000,	90.000,	0.000	W16E1	72.000,	90.000,	0.000	8.75E-01	2
16	W15E2	72.000,	90.000,	0.000	W17E1	116.000,	90.000,	0.000	7.50E-01	4
17	W16E2	116.000,	90.000,	0.000	W18E1	152.000,	90.000,	0.000	6.25E-01	4
18	W17E2	152.000,	90.000,	0.000		210.700,	90.000,	0.000	5.00E-01	6
19		-200.80,	139.520,	0.000	W20E1	-152.00,	139.520,	0.000	5.00E-01	5
20	W19E2	-152.00,	139.520,	0.000	W21E1	-116.00,	139.520,	0.000	6.25E-01	4
21	W20E2	-116.00,	139.520,	0.000	W22E1	-72.000,	139.520,	0.000	7.50E-01	4
22	W21E2	-72.000,	139.520,	0.000	W23E1	-48.000,	139.520,	0.000	8.75E-01	2
23	W22E2	-48.000,	139.520,	0.000	W24E1	48.000,	139.520,	0.000	1.00E+00	9
24	W23E2	48.000,	139.520,	0.000	W25E1	72.000,	139.520,	0.000	8.75E-01	2
25	W24E2	72.000,	139.520,	0.000	W26E1	116.000,	139.520,	0.000	7.50E-01	4
26	W25E2	116.000,	139.520,	0.000	W27E1	152.000,	139.520,	0.000	6.25E-01	4
27	W26E2	152.000,	139.520,	0.000		200.800,	139.520,	0.000	5.00E-01	5
28		-194.62,	266.700,	0.000	W29E1	-152.00,	266.700,	0.000	5.00E-01	4
29	W28E2	-152.00,	266.700,	0.000	W30E1	-116.00,	266.700,	0.000	6.25E-01	4
30	W29E2	-116.00,	266.700,	0.000	W31E1	-72.000,	266.700,	0.000	7.50E-01	4
31	W30E2	-72.000,	266.700,	0.000	W32E1	-48.000,	266.700,	0.000	8.75E-01	2
32	W31E2	-48.000,	266.700,	0.000	W33E1	48.000,	266.700,	0.000	1.00E+00	9
33	W32E2	48.000,	266.700,	0.000	W34E1	72.000,	266.700,	0.000	8.75E-01	2
34	W33E2	72.000,	266.700,	0.000	W35E1	116.000,	266.700,	0.000	7.50E-01	4
35	W34E2	116.000,	266.700,	0.000	W36E1	152.000,	266.700,	0.000	6.25E-01	4
36	W35E2	152.000,	266.700,	0.000		194.620,	266.700,	0.000	5.00E-01	4
37		-194.63,	388.440,	0.000	W38E1	-152.00,	388.440,	0.000	5.00E-01	4
38	W37E2	-152.00,	388.440,	0.000	W39E1	-116.00,	388.440,	0.000	6.25E-01	4
39	W38E2	-116.00,	388.440,	0.000	W40E1	-72.000,	388.440,	0.000	7.50E-01	4
40	W39E2	-72.000,	388.440,	0.000	W41E1	-48.000,	388.440,	0.000	8.75E-01	2
41	W40E2	-48.000,	388.440,	0.000	W42E1	48.000,	388.440,	0.000	1.00E+00	9
42	W41E2	48.000,	388.440,	0.000	W43E1	72.000,	388.440,	0.000	8.75E-01	2
43	W42E2	72.000,	388.440,	0.000	W44E1	116.000,	388.440,	0.000	7.50E-01	4
44	W43E2	116.000,	388.440,	0.000	W45E1	152.000,	388.440,	0.000	6.25E-01	4
45	W44E2	152.000,	388.440,	0.000		194.630,	388.440,	0.000	5.00E-01	4
46		-187.39,	570.000,	0.000	W47E1	-152.00,	570.000,	0.000	5.00E-01	4
47	W46E2	-152.00,	570.000,	0.000	W48E1	-116.00,	570.000,	0.000	6.25E-01	4

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48   W47E2 -116.00,570.000,  0.000 W49E1 -72.000,570.000,  0.000 7.50E-01   4
49   W48E2 -72.000,570.000,  0.000 W50E1 -48.000,570.000,  0.000 8.75E-01   2
50   W49E2 -48.000,570.000,  0.000 W51E1  48.000,570.000,  0.000 1.00E+00   9
51   W50E2  48.000,570.000,  0.000 W52E1  72.000,570.000,  0.000 8.75E-01   2
52   W51E2  72.000,570.000,  0.000 W53E1 116.000,570.000,  0.000 7.50E-01   4
53   W52E2 116.000,570.000,  0.000 W54E1 152.000,570.000,  0.000 6.25E-01   4
54   W53E2 152.000,570.000,  0.000       187.390,570.000,  0.000 5.00E-01   4

             ------------- SOURCES -------------

Source    Wire      Wire #/Pct From End 1    Ampl.(V, A)  Phase(Deg.)  Type
          Seg.     Actual      (Specified)

1           5    14 / 50.00   ( 14 / 50.00)      1.000       0.000       V

No loads specified

No transmission lines specified

Ground type is Free Space
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                     EZNEC/4  ver. 2.0

Hy-Gain 205-XLB, 48' Boom (WB0DGF)                07-19-1998     07:55:56


Frequency = 14.175  MHz.

Wire Loss: Aluminum -- Resistivity = 4E-08 ohm-m, Rel. Perm. = 1


              --------------- WIRES ---------------

Wire Conn. --- End 1 (x,y,z : in)  Conn. --- End 2 (x,y,z : in)   Dia(in) Segs

1          -286.00,-223.75,  0.000  W2E1 -286.00,-166.00,  0.000 4.38E-01   7
2     W1E2 -286.00,-166.00,  0.000  W3E1 -286.00,-142.00,  0.000 6.25E-01   3
3     W2E2 -286.00,-142.00,  0.000  W4E1 -286.00,-91.500,  0.000 8.75E-01   6
4     W3E2 -286.00,-91.500,  0.000  W5E1 -286.00,-45.500,  0.000 1.12E+00   6
5     W4E2 -286.00,-45.500,  0.000  W6E1 -286.00, -3.625,  0.000 1.25E+00   5
6     W5E2 -286.00, -3.625,  0.000  W7E1 -286.00,  3.625,  0.000 4.00E+00   1
7     W6E2 -286.00,  3.625,  0.000  W8E1 -286.00, 45.500,  0.000 1.25E+00   5
8     W7E2 -286.00, 45.500,  0.000  W9E1 -286.00, 91.500,  0.000 1.12E+00   6
9     W8E2 -286.00, 91.500,  0.000 W10E1 -286.00,142.000,  0.000 8.75E-01   6
10    W9E2 -286.00,142.000,  0.000 W11E1 -286.00,166.000,  0.000 6.25E-01   3
11   W10E2 -286.00,166.000,  0.000       -286.00,223.750,  0.000 4.38E-01   7
12         -170.00,-209.50,  0.000 W13E1 -170.00,-156.50,  0.000 4.38E-01   7
13   W12E2 -170.00,-156.50,  0.000 W14E1 -170.00,-132.50,  0.000 6.25E-01   3
14   W13E2 -170.00,-132.50,  0.000 W15E1 -170.00,-82.000,  0.000 8.75E-01   6
15   W14E2 -170.00,-82.000,  0.000 W16E1 -170.00,-36.000,  0.000 1.12E+00   6
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16	W15E2	-170.00,-36.000,	0.000	W17E1	-170.00,-10.000,	0.000	1.25E+00	3
17	W16E2	-170.00,-10.000,	0.000	W18E1	-170.00, 10.000,	0.000	1.25E+00	3
18	W17E2	-170.00, 10.000,	0.000	W19E1	-170.00, 36.000,	0.000	1.25E+00	3
19	W18E2	-170.00, 36.000,	0.000	W20E1	-170.00, 82.000,	0.000	1.12E+00	6
20	W19E2	-170.00, 82.000,	0.000	W21E1	-170.00,132.500,	0.000	8.75E-01	6
21	W20E2	-170.00,132.500,	0.000	W22E1	-170.00,156.500,	0.000	6.25E-01	3
22	W21E2	-170.00,156.500,	0.000		-170.00,209.500,	0.000	4.38E-01	7
23		-80.000,-206.25,	0.000	W24E1	-80.000,-156.50,	0.000	4.38E-01	6
24	W23E2	-80.000,-156.50,	0.000	W25E1	-80.000,-132.50,	0.000	6.25E-01	3
25	W24E2	-80.000,-132.50,	0.000	W26E1	-80.000,-82.000,	0.000	8.75E-01	6
26	W25E2	-80.000,-82.000,	0.000	W27E1	-80.000,-36.000,	0.000	1.12E+00	6
27	W26E2	-80.000,-36.000,	0.000	W28E1	-80.000, -3.625,	0.000	1.25E+00	4
28	W27E2	-80.000, -3.625,	0.000	W29E1	-80.000, 3.625,	0.000	4.00E+00	1
29	W28E2	-80.000, 3.625,	0.000	W30E1	-80.000, 36.000,	0.000	1.25E+00	4
30	W29E2	-80.000, 36.000,	0.000	W31E1	-80.000, 82.000,	0.000	1.12E+00	6
31	W30E2	-80.000, 82.000,	0.000	W32E1	-80.000,132.500,	0.000	8.75E-01	6
32	W31E2	-80.000,132.500,	0.000	W33E1	-80.000,156.500,	0.000	6.25E-01	3
33	W32E2	-80.000,156.500,	0.000		-80.000,206.250,	0.000	4.38E-01	6
34		86.000,-203.25,	0.000	W35E1	86.000,-156.50,	0.000	4.38E-01	6
35	W34E2	86.000,-156.50,	0.000	W36E1	86.000,-132.50,	0.000	6.25E-01	3
36	W35E2	86.000,-132.50,	0.000	W37E1	86.000,-82.000,	0.000	8.75E-01	6
37	W36E2	86.000,-82.000,	0.000	W38E1	86.000,-36.000,	0.000	1.12E+00	6
38	W37E2	86.000,-36.000,	0.000	W39E1	86.000, -3.625,	0.000	1.25E+00	4
39	W38E2	86.000, -3.625,	0.000	W40E1	86.000, 3.625,	0.000	4.00E+00	1
40	W39E2	86.000, 3.625,	0.000	W41E1	86.000, 36.000,	0.000	1.25E+00	4
41	W40E2	86.000, 36.000,	0.000	W42E1	86.000, 82.000,	0.000	1.12E+00	6
42	W41E2	86.000, 82.000,	0.000	W43E1	86.000,132.500,	0.000	8.75E-01	6
43	W42E2	86.000,132.500,	0.000	W44E1	86.000,156.500,	0.000	6.25E-01	3
44	W43E2	86.000,156.500,	0.000		86.000,203.250,	0.000	4.38E-01	6
45		286.000,-193.12,	0.000	W46E1	286.000,-138.25,	0.000	4.38E-01	7
46	W45E2	286.000,-138.25,	0.000	W47E1	286.000,-114.25,	0.000	6.25E-01	3
47	W46E2	286.000,-114.25,	0.000	W48E1	286.000,-63.750,	0.000	8.75E-01	6
48	W47E2	286.000,-63.750,	0.000	W49E1	286.000,-17.750,	0.000	1.12E+00	6
49	W48E2	286.000,-17.750,	0.000	W50E1	286.000, -3.625,	0.000	1.25E+00	2
50	W49E2	286.000, -3.625,	0.000	W51E1	286.000, 3.625,	0.000	4.00E+00	1
51	W50E2	286.000, 3.625,	0.000	W52E1	286.000, 17.750,	0.000	1.25E+00	2
52	W51E2	286.000, 17.750,	0.000	W53E1	286.000, 63.750,	0.000	1.12E+00	6
53	W52E2	286.000, 63.750,	0.000	W54E1	286.000,114.250,	0.000	8.75E-01	6
54	W53E2	286.000,114.250,	0.000	W55E1	286.000,138.250,	0.000	6.25E-01	3
55	W54E2	286.000,138.250,	0.000		286.000,193.125,	0.000	4.38E-01	7

---------- SOURCES ----------

Source	Wire Seg.	Wire #/Pct From End 1 Actual	(Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	2	17 / 50.00	(17 / 50.00)	0.707	0.000	V

No loads specified

No transmission lines specified

```
Ground type is Free Space
```

Any transcription errors in the above models are strictly my own fault, and I shall gladly correct any found.

Updated 7-20-98. © L. B. Cebik, W4RNL. Data may be used for personal purposes, but may not be reproduced for publication in print or any other medium without permission of the author.

Go to Amateur Radio Page

Projects and Info Center

Check out The Optimized Wideband Antenna (OWA) Yagi article by my very good friends Jim Breakall (WA3FET) and Nathan Miller (NW3Z). Jim and Nathan have also designed the SKYHAWK. A State-of-the-Art Trapless Tribander from Bencher/Butternut.

An Upgraded version of the OWA article and pics of the WP3R Contest site owned by WA3FET

- OWA
- UPGRADED OWA ARTICLE
- SKYHAWK
- RACZPR (Radio Astronomy Coordination Zone in Puerto Rico [Spanish]).

Back to KP4AO HOME page.

---->Please send any comments or suggestions to Angel M. Vazquez - WP3R angel@naic.edu.

Last updated Sunday, February 15, 1998

[Refresh this Menu]

[TEARA]

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- Club Bulletins
- Calendar
- Club Stuff
- The Repeaters
- FCC/License Info
- TEARA's Links
- APRS WX Info

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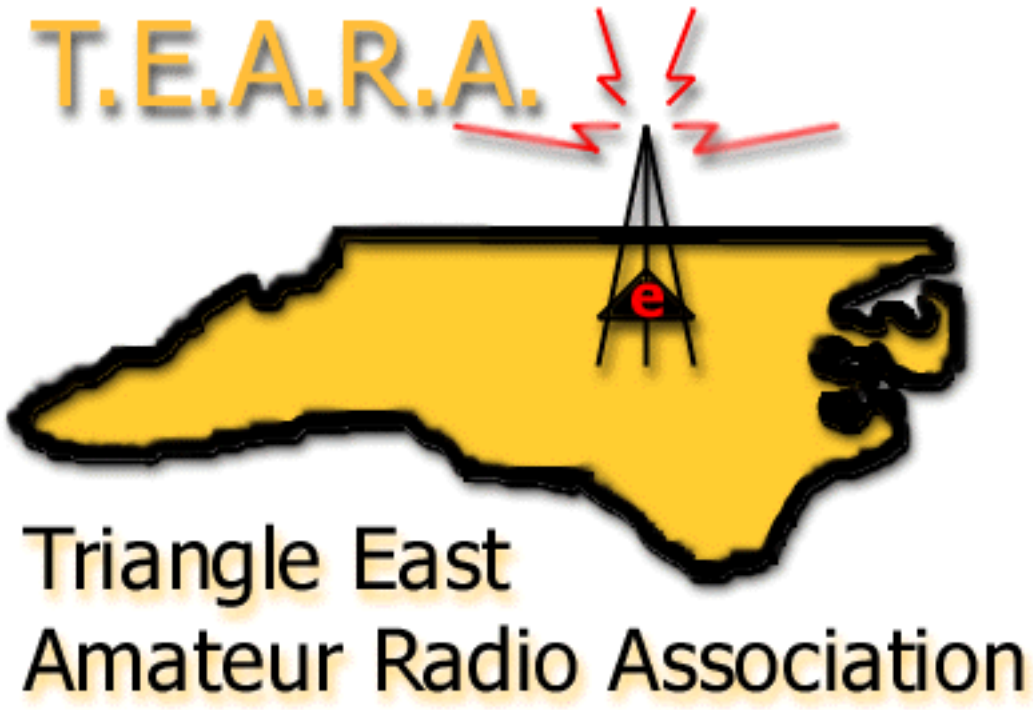
- Shops/Dealers
- Ham News
- Packet Radio
- ATV!
- Knowledge Quest
- Ham Classifieds
- Software

US<-Find Calls->DX |

Scoot Works Custom Motorcycle Accessories |



TEARA - The Triangle East ARA

3509 Rolesville Road; Wendell, NC 27591-9135

WAIT!! **If you came to this page directly from a search engine or link, and don't see a menu to the left and top of this page, you're missing out on hundreds of pages of amateur related info on our server. CLICK HERE, or click the large NC State Logo at the top, to move to the top of our site (which automatically loads this page). From there, you can select hundreds of pages of information on our server, complete with all of our site navigation tools. Enjoy your visit with TEARA!**

The TEARA Email Reflector

An Email Group for Triangle area Amateurs!

The TEARA email list is primarily for TEARA (Triangle East Amateur Radio Association) members, and those interested in membership in central NC. It is specific to this local amateur radio club for the purpose of distributing club bulletins and promoting communication within the club membership.

The list can be subscribed to as a normal list, or as a digest. Here is the subscription info:

- **To subscribe to the list through email, send an email to: teara-subscribe@yahoogroups.com**
- **To unsubscribe to the list through email, send an email to: teara-unsubscribe@yahoogroups.com**
- **To switch your subscription to normal, send an email to: teara-normal@yahoogroups.com**
- **To switch your subscription to digest, send an email to: teara-digest@yahoogroups.com**

Once you're subscribed, send your messages to: teara@onelist.com and everyone who is subscribed will see them.

[Digest Version]

Now, having said that, I wonder how many of you are aware that this list is available in "DIGEST" format? The DIGEST format takes every day's email, rolls it into a single email, and sent it to you at once, rather than a bunch of individual emails. Personally, I prefer individual emails so I can easily see the subject and reply to them that way, but lots of people like the digest version for it's compactness. If you want to be set up on the digest version, drop me a note and I can make the change. You can also change to either normal or digest versions yourself by doing this:

- **To switch your subscription to digest, send a blank email (with nothing in the subject or message body, and no signature) to: teara-digest@yahoogroups.com**
- **To switch your subscription back to the normal delivery, send a**

blank email (with nothing in the subject or message body, and no signature) to:
teara-normal@yahoogroups.com

[TIPS for using the Digest Version]

- **1)- When replying to a digest message, be sure to take a look at the subject line. It'll usually state the digest # it's replying to, rather than the individual message within the digest. Change the subject to reflect what ever topic you're replying to, so folks will know what the message is about.**

- **2)- When you click "reply" to a Digest message, everything in the digest will be captioned or quoted as part of reply. Be sure to highlight and delete the rest of the digest that doesn't pertain to your specific reply, so the email (if you're clicking "reply") won't contain a bunch of other messages. That'll also help keep the email as small as practical.**

[ARCHIVES]

The TEARA list has mail archives of all the traffic on the list. These can be accessed with a web browser like Netscape or Internet Explorer. This could be helpful if you're looking for a missed note, or maybe at another terminal somewhere (public library, work, friend's house, etc) and want to check in to see what's going on. There is also an archive storing every digest of every message ever sent on the ncvulcan list.

- **To get to todays messages, click --> HERE<-- !**

I hope these tips improve or supplement your enjoyment of the TEARA list!

Webmaster: *webmaster@teara.org*

***RETURN* to the main TEARA Webpage!**

TRAMPING GROUND
Click HERE to continue!

Welcome to VE7CA's Antennas Page

Since becoming an amateur radio operator I have built a lot of antennas. I have had several articles published where I described antennas that have served me well. The above menu bar will take you to the articles.

December, 2003

One hour before the 2003 CQWW CW Contest, I decided to raise the telescoping tower that supports my modified Telerana LPDA, 30/40 meter dipole and 6 meter yagi. Just before it reached max height the tower collapsed into its self and came down with lightning speed. With the rapid deceleration and then sudden stop, the outer ends of the booms and fiberglass spider supports kept moving downward and broke a couple of the poles. The damage is evident in the photo below. This is a reminder to those who own crank up towers, check the cable, cable fasterners and winch brakes regularly!

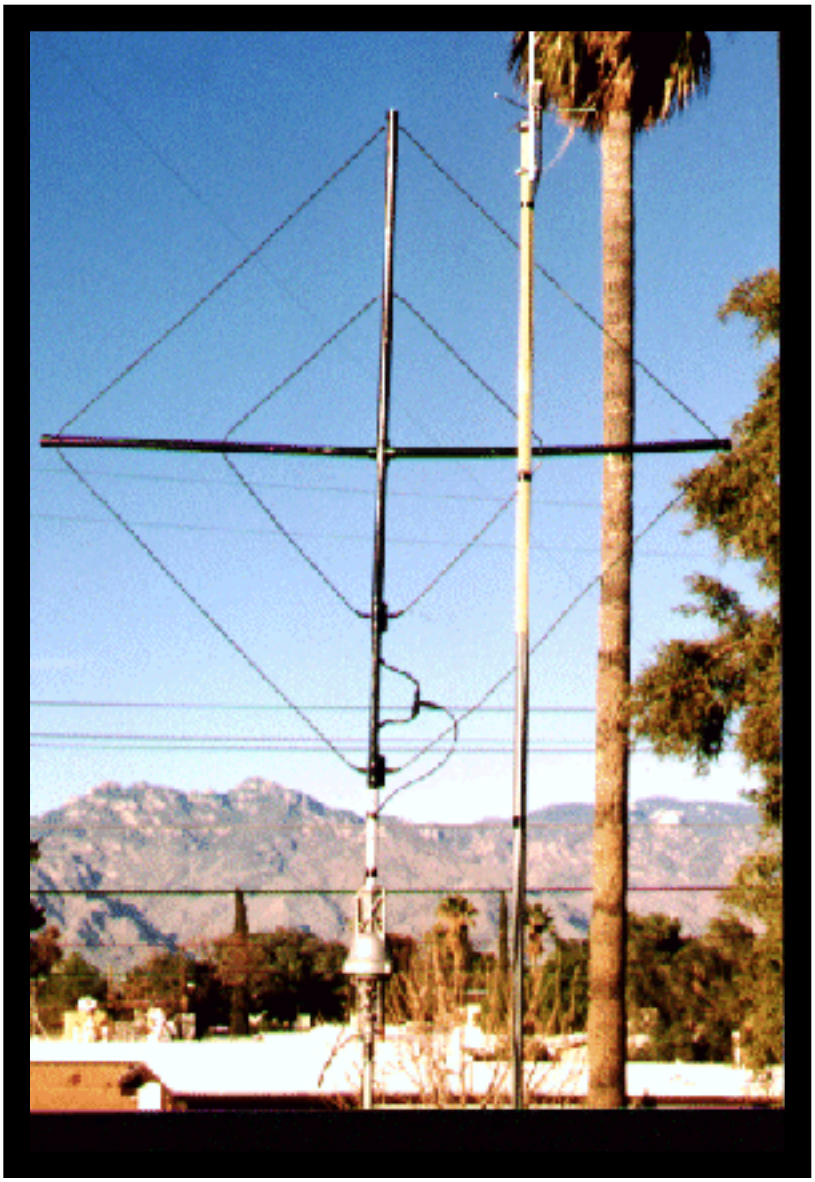
Fortunatly, no damage to VE7CA or the skylight through which I was watching the tower as I operated the switch that controls the electric winch.

August 23, 2004

After lowering the tower to determine why the tower collapsed, I found that the raising cable broke where it entered the electric winch. Why it broke I have no idea. There was also damage to the bottom ends of the each section where the winch cables are attached. I am in the process of repairing the damage and hope to have the tower up by the end of September 2004.

I am also in the process of designing a new antenna to replace the Modified Teleranna. Since I have been concentrating on QRP contesting of late, I have decided that I need an antenna with higher gain in the CW portions of 20, 15 and 10 meters. As well, gain on 40 meters would be much appreciated. I will keep you posted on my progress as I near a new design.

160/80m Coaxial Receiving Loops



NOTE: As of January 1, 2004 I must raise the price of my receive loops. As that my construction costs, have risen.

Several years ago a friend of mine (W7AE) and myself became interested in low band dxing. Lacking room for a beverage receiving antenna, we constructed several loop receiving antenna to see if we could find a suitable low /lower noise receiving antenna.

W7AE built the first loop for 160m following plans laid out in the ARRL Antenna Handbook. Encouraged by the results I built one for 80m. I found that with a proper pre-amp, such the Palomar series, or Ameco's PT-3 there was a significant lowering of noise, but the signals that I was unable to hear on my sloper now were quite workable using the receiving loop. For the past several years, we experimented with several different configurations , including a circular, and square loop. We found that the diamond configuration worked the best.The 80m version requires a pre-amp to bring the signal up to an acceptable quality. We've found that there is a one to two s unit difference (depending how high you set the gain on the preamp)in the lowering of noise. There have been several times that the loop has made the difference between working a DX station or not hearing him on my sloper.Though the antenna is not a beverage, we've found it to be a good alternative.

Click Here If you want me to construct one for you.

Click here If you wish to build one yourself.

Low Band Links (Presently under reconstruction.)

The latest news, is that now I can accept Paypal. Click on the link that says "If you want me to construct one for you" for more details.

Please note that there is a new e-mail address and price increase on the 160/80m loop. (Component prices seem to be going up.)

Your are visitor number

updated 08/22/2001

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Lancaster Amateur Radio Club

Ham Radio Station W2SO

ARRL Amateur Radio

Antenna's "How To" Build Your Own

Ham Gear For Sale

Weather Station

Radio Nets WNY

News Letter

Club Calendar

About Ham Radio

Shortwave Radio

Dunkirk Lighthouse

Blizzard of '77

PHOTO'S

Super Linear - Loaded Inverted V

[Home] [Our Members] [Our Club] [Greater Buffalo Hamfest] [Ham Radio Information] [Ham Radio News] [Contact US]

[Standard Dipole Construction] [Super Loop Antenna] [J - Pole] [Pocket J - Pole] [Quickie Vertical]
[Super Linear - Loaded Inverted V]

The KGØZP Super Linear-Loaded Inverted V

How do you fit a full length 160 meter antenna into a 40 foot deep yard?
Install the KGØZP Super Linear-Loaded Inverted V, of course!

The Super Linear-Loaded Inverted V allows you to select your two favorite and most used frequencies as band centers, and offers excellent bandwidth as well.

The designers installation centerpoints are set at 1.840 the center of
the DX window or calling frequency and 1.875 at an SWR of 1.1-1
The bandspread at 1.5-1 is 20 kc at each center or
40 kc to an SWR of 2-1

The linear-loading section is mounted near enough to the ground that
you can easily install or remove precut pigtails using an alligator clip
to move the center point temporarily to make that sked, or you
can tune the whole 160 meter band with a simple antenna tuner.

The formula for construction is the same as for any dipole 468/fMhz
and the design will work equally well on 80 or 40 meters.

Reader Please Note: The photos on this page are greatly compressed to insure quick downloading.
The link accompanying each photo will cause the uncompressed full size photo to download.

It would be impossible to show an overall view of the whole antenna system in such a way that you could see the wires.

With that in mind, I have taken a couple of daylight pictures as well as flash night pictures which allowed me to highlight the wires against a black sky.

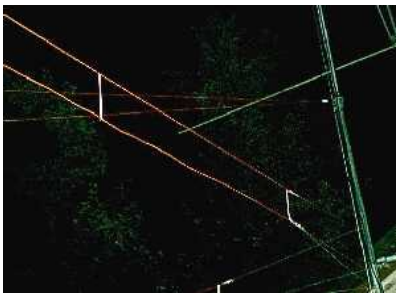
Lets take a thumbnail look at a daylight view of the Rohn 50 foot push-up pole utilized for this installation.

This is quite a busy photograph, so I will name the items in descending order to help pinpoint the areas relavent to the Super Linear-Loaded Inverted V.

The white stick at the top extending off the upper edge of the photo is a Diamond Tri-bander, model X3200A, fed with 9913 coax.
Directly below the tribander is the first guying ring. This is a dummy ring is used to ground and hold the upper wire of a Wermager type Broadband Sloper. In the full sized photo you can just barely see this wire extending left forward.
The highest of the three actual guying rings (second one down from the top) is at the 40 foot high point on the tower, a pulley is connected to this guying ring and through it is threaded a high quality flagpole rope. This rope can be used to raise and lower the balun for experimenting with different antenna designs, without changing the position of the push-pole.
The Super Linear-Loaded Inverted V, the topic of discussion is connected to this balun. The tension on the wires is so great that it holds the balun outward, the line descending from the bottom of the balun is Super Mini-8 coax. The pull of the Inverted V's two antenna wires is offset by two guywires on the other side of the guying ring. I have a wide yard and could keep the apex slightly wider than 90 degrees.
The next guy ring down is at the 30 foot height point and has nothing more three guy wires.
The lowest guy ring visible in the photograph is at the 20 foot level, and is the feedpoint for the Fan portion of a Wermager type Broadband Sloper originally used on 160 but now cut for 80 meters. I would like to mention again that the upper wire of the Wermager type Broadband Sloper is the wire grounded to the push-pole at the dummy guy ring at the very top. The other end of this wire is tied to the same location as the guy wire coming from the 30 foot guy ring, which places the metal guy wire directly between the driven Fan and the grounded sloper element of the Wermager type Broadband Dipole.

This photo is a view showing the East outbound leg of the Super Linear-Loaded Inverted V. In the upper left of the photo, you can see the single sloping wire descending from the balun. The two horizontal wires below comprise part of the Linear-Loading section.

The remaining photo's were taken at night using a flash, then enhanced to make the wires more prominent.



In this first photo, you can easily see that the Linear-Loading section is a good distance away from the push-pole. I used a distance of 5 feet in this installation, but you can reduce this distance to 3 feet if necessary. The vertical spacers are drinking straws, one nested inside the other for greater strength, a nylon tie line is fed through the center of the straws.
The upper tie off lines are located 1 foot above the standoff mounted to the eave. The lower tie off lines should also be located at or just above the standoff. Original twine used during initial setup is still visible and should have been removed before taking the photo.

The white insulator some 5 feet above the linear loading section is the feedpoint for the Fan portion of the Wermager type Broadband Sloper.

SPECIAL NOTATION: The upper wire and the lower wire of the linear-loading section are connected at this end only. This will be described in greater detail in the construction notes section.

These last two photos show the East and West outbound ends of the Super Linear-Loaded Inverted V.

If you look closely, you can see the pigtails wrapped around the lower nylon guys.

In the photo on the left of the East outbound end, the wire from the balun down to the insulator is not visible. The upper guy wire from the insulator is metal, the lower is nylon. An additional twisted wire above the insulator is used to lift the linear loading section slightly higher than the existing guy mount allows, I did this only because I had something above each outbound end to tie to.

The right photo shows slightly greater detail and the antenna wire from the balun is clearly visible. There are no electrical connections at the outbound end of the linear-loading section.
Again, if you look closely, you can see the pigtail extension wrapped around the lower nylon guy.

Construction Notes:

The construction of the Super Linear-Loaded Inverted V is quite simple and follows the same formula as for a standard dipole. 468/fMhz (FourHundredSixtyEight divided by the Frequency in Megahertz give you the total length in feet for a 1/2 wave dipole). In practice you will use two wires, each being 130 feet long. The antenna shown in the photos is made from 14 guage insulated copper wire. I ran out of insulators and am temporarily using loops made from 1/4 inch nylon cord at the push-pole end of the linear-loading section. If you start with two 130 foot long wires, you do not have to measure the turning or ending points, they will fall in place depending upon the height you install the horizontal portion of the antenna.

The height of the feedpoint can vary considerably, however, the higher the better. If you can only go up 30 or 35 feet, the antenna will still perform almost as well, but you will require a further horizontal distance to work in. With the feedpoint at 40 feet, the sloping elements are roughly 60 feet long at 40 feet away from the push-up pole. Any distance from 40 feet to 80 feet in length seems to make little difference in the performance of this antenna.

A balun is not an absolute necessity, however, if space constraints keep you from obtaining a spread between the sloping wires greater than 90 degrees, then a balun is strongly suggested.

For all practical purposes, installation is the same as for an inverted V, a feed type insulator or balun is assembled with the coax and both of the 130 foot long wires. Connection of the coax is the same as for a standard dipole. The center conductor goes to one wire and the shield to the other wire. The feed insulator or balun is now hoisted to it's permanent position on your push-pole, tower, gable or wherever you are mounting your antenna.

I will describe the finished setup first, then give some quick install tips immediately following.

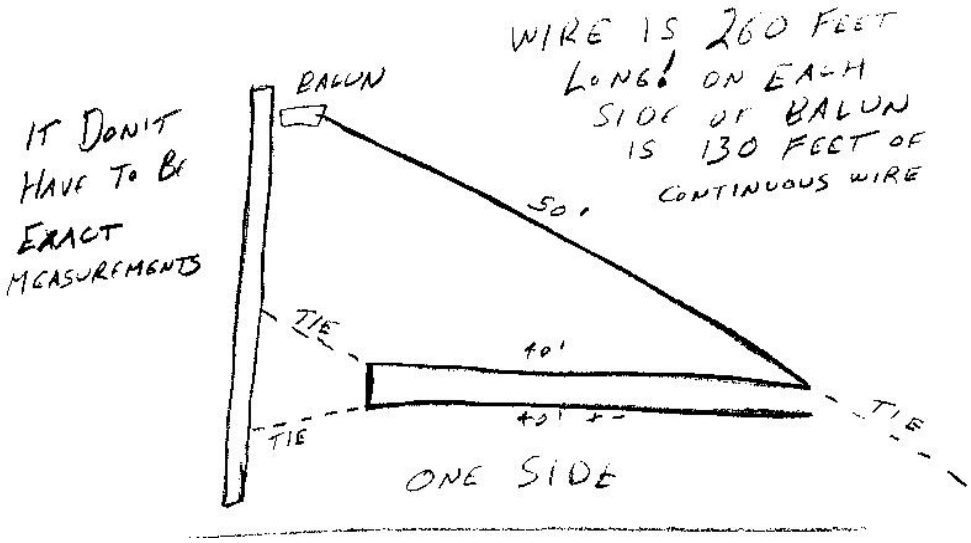
One side of the antenna is assembled first and will require a minimum of three insulators or you can use nylon mounts by giving the wire a twist to form a loop to connect the nylon rope to.
The antenna wire is brought out across the yard and the end of the wire is passed through an insulator at the outbound end of the antenna. The antenna wire is then brought back toward the push-pole and passed through the second insulator. The second insulator should be installed about 6 feet away from the push-pole temporarily. The wire then passes through a third insulator 6 to 8 inches below the second insulator, and then proceeds back out to the outbound end.

The ideal installation would have the feedpoint 50 feet above ground level and the linear-loading section 20 feet above ground level. In practice, the feedpoint can be as low as 30 feet above ground level and the linear-loading section as low as 10 feet from the ground without affecting performance to drastically. In any case, the linear-loading section should be maintained as close to horizontal as possible.

The idea is to determine at what height you wish to install the horizontal linear-loading section, and then adjust the insulators and distance between the push-pole and inside end of the linear-loading section so that the first insulator and end of the wire are on the same plane. And you'll really love this, give or take 2 or 3 feet on that tail, preferably give a few feet extra for trimming to resonance.

Any non-metal spacer can be used to maintain distance between the elements of the linear-loading section! No need to go overboard here, it's close enough to the ground you can replace the simple item I use in Quick Tips below.

Simple Sketch showing path of antenna wire on one side of balun.



Quick Tips:

Here is how to install this antenna system and have it up and running in well under an hour, exluding push-pole installation of course.

1. Install the wires to your balun and connect the coax, hoist it into position and secure firmly.
2. Pull one wire out taught to your guy mount and secure temporarily.
3. Take the loose end of this same wire and secure it to the push-pole temporarily at the desired final installation height. Note: This end of the wire will be at the other end of the yard when finished.
4. Go back to the outbound end of the antenna, using a stepladder if necessary, pull the wire so that it is taught from the push-pole and lift the wire until it is perfectly horizontal.
5. With a magic marker or piece of tape, mark the sloping wire at the point where the wire from the push-pole crosses the sloping wire. This is where the outbound insulator is to be installed.
6. Install the outbound insulator and tie off (guy) this portion of the antenna to permanent tension.
7. Disconnect the loose end of the wire from the push-pole, pass it through an insulator and bring the loose end out to the outbound end and temporarily connect it to the now secure insulator outbound insulator, leaving 2 to 3 feet of loose pigtail extending through this insulator.
8. Pull the wires from the insulator toward the push-pole until they are in balance at equal lengths from the outbound end, at this bend take the insulator and give it a twist, release the loose pigtail at the outbound end, then secure the push-pole end of the top linear-loading wire and it's insulator to the push-pole at it's permanent position.
9. Install your first spacer (see spacers below) by passing the loose end of the wire through it, bring it up snug to the insulator, install the second insulator and give it a twist to hold it secure. Tie this insulator off to the push-pole as well, trying to keep the spacer vertical.
10. Install additional spacers, one about every five feet, to maintain separation of the upper and lower linear-loading sections.
11. When you reach the outbound end, install your last spacer at the insulator, form a loop in the lower wire for attaching a nylon tie off rope. Pull the rope to tighten the lower linear-loading wire and tie off to the guy mount. Wrap the remaining 2 or 3 feet of pigtail around the nylon rope.
12. Duplicate the above instructions for the other leg of the antenna!

SPACERS:
I take simple drinking straws to use for spacers. But reinforce them to two thicknesses for all the spacers except the inbound spacer near the push pole, this one is three or four thicknesses.
To strengthen a soda straw, you merely slit another soda straw end to end and insert it inside of the whole unslit soda straw. You only need to pass the unslit whole soda straw over the wire for the inbound spacer, the other two or three slit straws can be added right over the wire and then slid into the whole unslit straw.

Joining the straws to the wires:
I take 24 inch long pieces of braided nylon, like mini-blind sash cord, fold it in half, hold it on one side of the top wire with the loop up and open, then I pass the loose ends of the rope over the wire and through the loop, pull it snug and let the ropes hang until I get all the ties (ropes) installed.
By taking a 12 inch long piece of 18 or 20 guage wire and bending a sharp turn at one end, you can use that to slip the ends of the ropes into and slide the soda straw over the wire pulling the rope down through the straw. Place the loose ends of the rope one on each side of the lower wire and tie a square knot, drawing the wire up tight against the soda straws. Don't worry if they are atilt right now, you can true them up later.
After the antenna is assembled, tied down nice and snug, and tuned. I will adjust the soda straws to vertical and place a dab of something like plasti-dip-your-grip or dumb gum over the tops of the straws to hid the nylon braid from UV and weather. I usually don't worry about the knot end or lower end of the straw.

Tune-Up:
If I have obstructions in the yard that interfere more with one leg than the other, I will make the shield side of the coax go to that side and the center conductor to the clearer of the legs. Now I tell you!
I normally tune both legs of the antenna first to 1.840 MHz, then I will continue to cut the clearer leg of the antenna all the way down to 1.950 MHz. I will then make a pigtail with an alligator clip on it and retune that leg of the antenna to 1.875 MHz. You can wrap the pigtail around the nylon guy and slide it down at least 6 inches away from the tuned pigtail end and it won't interfere with your upper band setting. In my case, I leave the pigtail connected as 1.875 MHz is one of my popular areas of the band for ragchewing. I may be over 50, but to really fit in up at 1.950, I need another 40 years under my belt, Hi Hi..... You may make other pigtails to center in other areas of the band also. It works on either leg of the antenna!

TTUL - 73+ de Gary - KGØZP

Questions or comments may be e-mailed directly from the home (index) page.

© 1999 Gary V. Deutschmann, Sr. - SCS/RAIAR

Amateurfunk Ham Radio Antennas

Operator: Norbert Prenzel (DJ9RB - VU2RBQ - H44 / DJ9RB)

Short Dipoles and Verticals

Verkürzte Antennen (160m, 80m, 40m)

160m Band, Loaded Dipole	Dipol mit Endkapazität
160m Band, Loaded L-ANT (Pattern)	L-Antenne mit Endkapazität
160m Band, Kite with Vertical Antenna	Drachen mit Vertikalantenne full size
80m Band, Short Dipole	kurzer Dipol für das 80m Band
80m Band, Top Loaded Vertical	Vertikalantenne mit Dachlast, 8m hoch, f = 3,5 MHz
80m Band, Top Loaded Vertical	using elevated radials - Radials 1m über Grund new
80m Band, Shielded Loop	abgeschirmte Loop für 3,5 MHz
40m Band, Top Loaded Vertical	Vertikalantenne mit Dachlast, 7m hoch, f = 7 MHz
40m Band, portable with Top Load	Portabel-Antenne mit Dachlast, 6m hoch, f = 7 MHz
40m Band, portable Dipole	Portabel-Dipol, f = 7 MHz
40m Band, Loop-magnetic	Loop (verkürzte Rahmenantenne), f = 7 MHz (80,40,30 m)
Antenna Measure Instruments	Eigenbau Meßgeräte

Pictures

Fotos

VU2*India - Location (QTH)	80m Band Vertical on Top of the Hotel
VU2*India - TOP LOAD	Sphere from the 80m Band Vertical
Radio Operator	Norbert DJ9RB (VU2RBQ)
H44*Solomon Isl - QTH	at Faisi Island
H44*Solomon Isl - Equipment	is running from Battery
H44*Solomon Isl - Equipment	Shack (Funkraum)
H44*Solomon Isl - OPERATORS	H44 / DJ9RB , H44MS (DL2GAC)

Activity

Aktivitäten

INDIA Log 160m Log 1996-2004

Fly a Kite with Antenna adapted

Log from VU2RBQ (VU2 / DJ9RB)

Drachen mit Antenne im 160m Band

Nostalgia

Wiring Diagram and Pictures

Mein erster Sender

Schaltplan meines Eigenbausenders aus dem Jahr 1964 und Fotos new

Links

VU2RBQ Norbert

DL0PW OV-Würmsee (Starnberger See)

DL2UV and other Links

DF3CB Homepage

DX

HF-Cluster DX-Info 160m ... 10m

DX-Info QSL.Net

more Links

Propagation

Solar Terrestrial Report

MUF-Map Propagation WWV

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dj9rb @ arcor . de

28.03.2001

28.03.2001

Start was at 24.10.1996

Last change on: Aug.2004

Half-Length 80-Meter Vertical Monopoles: the Best Method of Loading

L. B. Cebik, W4RNL

Summary

In the pursuit of obtaining the most compact and efficient 80- meter monopole antenna, numerous loading schemes have been proposed, including based lumped constant loading, base linear loading, top (capacity) hat loading, and a number of antenna element extension-and-fold-back systems. Modeling these systems is difficult due to the various limitations of existing modeling software, including MININEC, NEC-2, and NEC-4.

Preliminary work is best done in MININEC, because (with due caution) it is best capable of modeling nonlinear geometries employing wires of different diameters, a necessary condition of a compact 80-meter monopole. A monopole 37.5' long, corresponding to a common commercial height, is the constant main element used, with other parameters varied to achieve the following goals:

- a. Maximum gain within the limits of the antenna type;
- b. True vertically-polarized circular pattern;
- c. Highest feedpoint impedance for maximum efficiency;
- d. Flattest SWR curve between 3.5 and 3.7 MHz; and
- e. Most compact and mechanically practical assembly.

The range of models compared covered the following types of monopoles: full-length, 37.5' unloaded, lumped constant base- loaded, linear-base-loaded, "capacity" hat loaded; top linear loaded; zig-zag fold-back loaded, and helically loaded. Figures are provided on gain and feedpoint impedance at 3.6 MHz, as well as on feedpoint impedance and SWR at 0.05 MHz intervals from 3.5 to 3.7 MHz. Other data can be obtained from these models by rerunning them using the descriptions provided in an Appendix. Especially recommended is a study of current levels along the antenna wires.

No single model antenna achieves all of the goals listed above. However, the helical fold-back element extension model achieves goals a. through c., and perhaps holds promise of achieving e. The zig-zag fold-back element extension model excels in achievement of goals b, c. and d., with only slightly less gain

than the helix and with some promise of meeting goal e. The capacity hat model shows excellent gan, bandwidth, and feedpoint impedance, but may be mechanically problematical except in a Marconi configuration. All other models show lesser performance in one or more categories. Unless mechanical constraints preclude further work on them, the helical and the zig-zag foldback models appear to be the best candidates for further study and testing. However, before hats are discarded as too large or too fixed or too unwieldy, the spiral hat noted in the last segment (Part 5) should be examined. It was modeled in NEC-4 because the model was too large of standard MININEC.

1. Goals, and Methods of the Study

2. Baseline Data: Full-Size and Capacity-Hat Verticals

3. Base-Loading: Lumped-Constant and Linear Loading

4. Top (Element-Extension) Loading: Linear, Zig-Zag, and Helical Loading

5. Summary Comparisons and Conclusions; With an Alternative Suggestion

6. Descriptions of Models Reported

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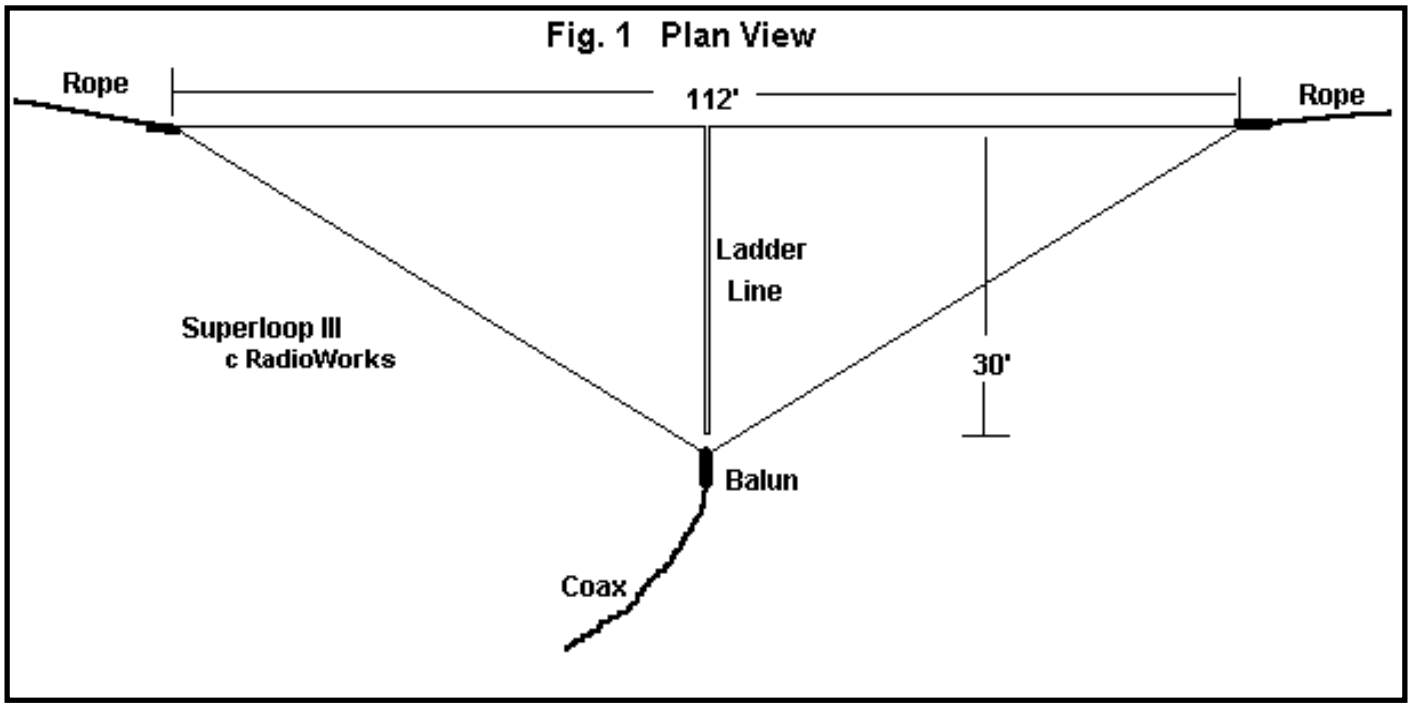
Lancaster Amateur Radio Club

Ham Radio Station W2SO

ARRL Amateur Radio

Super loop 80



G5RV verses Superloop 80

Many operators with small lots, a G5RV is what can fit for the 80 and 40 meter bands. The G5RV is 102 feet long and has a 34 foot
section of twinlead followed by coax into the shack, possibly with some sort of RF choke on the coax. The ends are typically supported by ropes up in
the trees. An 80 meter dipole would be about 134 feet long.

A tiny lot is limited in antenna potential and zoning laws prevent real towers.

RadioWorks "Superloop III" designed by Jim, W4FTU, and refined over the years, is a good alternative

PHYSICAL VARIATIONS

The standard arrangement is shown in Fig. 1. It looks like an inverted delta loop and is 112 feet across the top. It fit on the same ropes as my G5RV used and the coax even started at about the same point in space. The wire is heavy 14 gauge copper. If your space doesn't quite allow this, the top corner insulators can be moved to shorten the 112 foot dimension; also additional insulators can be added to the diagonal wires to make a rectangular

shape and raise the bottom balun up in the air more. I also added 6 feet of wire to move the resonant freq closer to the band bottoms for digital work.

The loop can also be mounted upside down and slanted if you only have a single support available. As with all loops, the area enclosed is important and so is the average height; the standard inverted delta shape is a very good compromise.

ELECTRICAL CHARACTERISTICS

The "trick" to the Superloop is the 30' length of ladder line hanging down from the center insulator. This length has been tuned so that appears to be a open-circuit stub on 40 meters; thus the antenna becomes two full-wave wires (at 40 meters) and is commonly referred to as the Bi-Square antenna. On 80 meters, it appears to be a short and the antenna becomes a single wave vertical loop. This happens automatically and no switching is involved.

A special balun is provided which gives a match between the 50 ohm coax lead-in and the higher resistance of the loop. For best matching, a 1/2 wavelength coax is recommended (e.g. 99' of RG-8X); however mine is about 70 feet into my diff-T tuner and the SWR < 2 points are 3495 to 3787 but the short coax gives a minimum on 40 of 2.05 at 7090 KHz. If you need to run without a tuner, close attention to the coax length will help. The balun is the typical ferrite rod in a PVC pipe with foaming urethane inside. This has the effect of heat insulating; mine works fine on 500 RTTY watts contesting, but real high power may be a problem on RTTY; but those guys all have beams, right?

OPERATING RESULTS

The diagonal wires make it partially a vertical antenna with a nice reduction in polarization QSB. You can possibly double contacts on 80/40 over the G5RV. RITTY can help on the reception. The Superloop tunes up fine on the 20,15,10 bands Antenna, ropes, and coax will run you about $US 135. RadioWorks advertises in CQ and QST and have an interesting catalog.

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By Joe Everhart, N2CX

The NJQRP Squirt

This reduced-size 80-meter antenna is designed for small building lots and portable use. It's a fine companion for the Warbler PSK31 transceiver.

At one time, 80 meters was one of the more highly populated amateur bands. Lately, it has become significantly less popular because much DXing has moved to the higher frequencies and many suburban lot sizes are too small to accommodate a full 130-foot, λ/2 antenna for the band. That's unfortunate, because 80 meters has lots of potential as a local-communication band—even at QRP levels. The recently published Warbler PSK31 transceiver can serve as a great facilitator for close-in QRP communication without much effort.[1] What's really needed to complement the Warbler for this purpose is an effective antenna that fits on a small suburban plot. Because PSK31 (which the Warbler uses) is reasonably effective even with weak signals, we can trade off some antenna efficiency for practicality.

What's a Ham to Do?

I investigated a number of antenna possibilities to come up with a practical solution. One intriguing candidate is the magnetic loop. Plenty of design information for this antenna is presented in *The ARRL Antenna Book* and at a number of Web sites.[2, 3] To obtain high efficiency, however, the loop must be 10 feet or more in diameter and built from 1/2-inch or larger-diameter copper pipe. The loop needs a very low-loss tuning capacitor and a means of carefully tuning it because of its inherently narrow bandwidth. Another configuration, the DCTL, may be a solution, but it's likely not very efficient.[4]

An old standby antenna I considered is the random-length wire worked against ground. If it is at least λ/4 long (a Marconi antenna) or longer, it can be reasonably efficient. Shorter lengths are likely to be several S units down in performance and almost any length end-fed wire needs a significant ground system to be effective. Of course, you may not need much of a ground with a λ/2 end-fed wire, but it's as long as a center-fed dipole.

Vertical antennas don't occupy much ground space, but suffer the same low efficiency as the end-fed wire if they are practical in size.

Probably the easiest antenna to use with good, predictable performance is the horizontal center-fed dipole. Unfortunately, as mentioned earlier, the usual 80-meter λ/2 dipole is too large for many lots. But all is not lost! The dipole can be reduced to about a quarter wavelength without much sacrifice in operation (see the sidebar, "Trade-Offs"). Furthermore, if the dipole's center is elevated and the ends lowered—resulting in an inverted **V**—it takes up even less room. This article describes just such a dipole: the NJQRP Squirt.

V for Victory

You can think of the Squirt as a 40-meter, λ/2 inverted-**V** dipole being used on 80meters. Figure 1 is an overall sketch of the antenna; Figure 2 is a photograph of a completed Squirt prior to erection. The Squirt has two legs about 34 feet long separated by 90° with a feed line running from the center. When installed, the center of the Squirt should be at least 20feet high, with the dipole ends tied off no lower than seven feet above ground. This low antenna height emphasizes high-angle NVIS (Near Vertical-Incidence Skyware) propagation that's ideal for 80-meter contacts ranging from next door out to 150 or 200 miles. And

Figure 2—An assembled Squirt ready for installation.

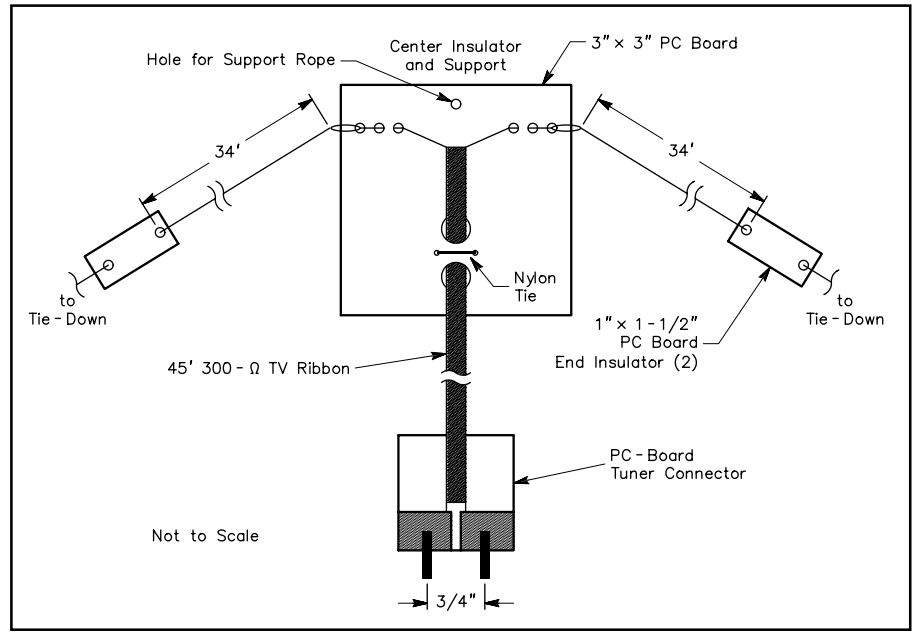

Figure 1—General construction of the 80-meter Squirt antenna.

[1]Notes appear on page 43.

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that's where 80 meters shines! With the Squirt's center at 30 feet and its ends at seven feet, the antenna's ground footprint is only about 50 feet wide.

One nice feature of a λ/2 center-fed dipole is that its center impedance is a good match for 50- or 75-Ω coax cable (and purists usually use a balun). Ah! But the Squirt is only λ/4 long on 80 meters, so it *isn't* resonant! Its feedpoint impedance is resistively low and reactively high. This means that feeding the antenna with coax cable would create a high SWR causing significant feed-line loss. To circumvent this, we can feed the antenna with a low-loss feed line and use an antenna tuner in the shack to match the antenna system to common 50-Ω coax cable. I'll have more to say about the tuner later.

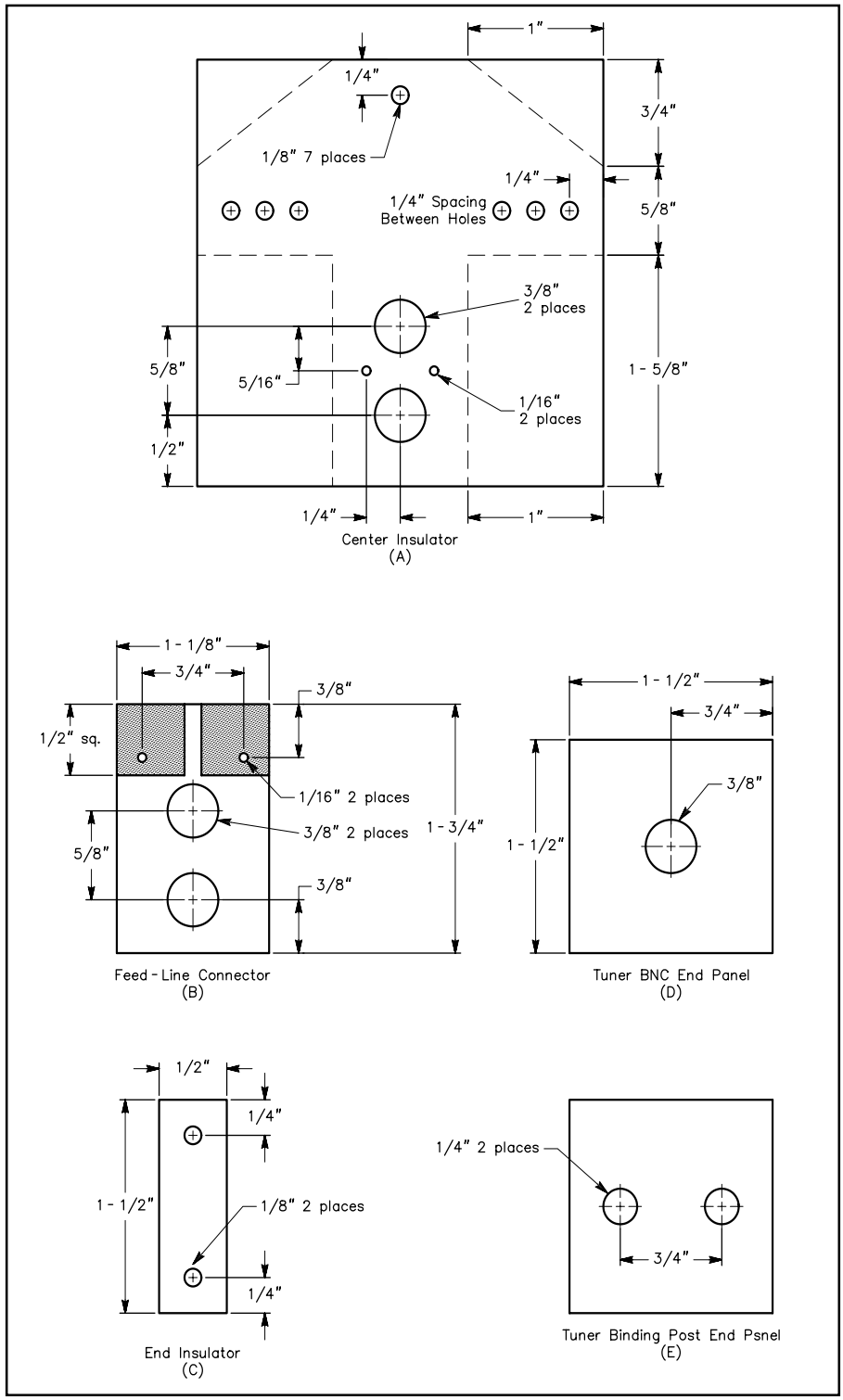

Figure 3—Hole sizes and locations for the various PC-board pieces. See Note 5.

I use 300-Ω TV flat ribbon line for the feed line. Although a better low-loss solution is to use open-wire line, that stuff is not as easy to bring into the house as is TV ribbon. Using TV ribbon sacrifices a little transmitted signal for increased convenience and availability. If you feel better using open-wire line, go for it!

Using Available Materials

It's always fun to see what you can do with junkbox stuff, and this antenna is one place to do it. See the "Parts List" for information on materials and sources.[5] For instance, the end and center insulators (see Figure 1) are made of 1/16-inch-thick scraps of glass-epoxy PC board. For the antenna elements, I use #20 or #22 insulated hookup wire. Although this wire size isn't recommended for use with fixed antennas, I find it entirely adequate for my Squirt. Because it's installed as an inverted **V** antenna, the center insulator supports most of the antenna's weight making the light-gauge wire all that's needed. The small-diameter wire has survived quite well for several years at N2CX. This is not to say, of course, that something stronger like #14 or #12 electrical house wire couldn't serve as well.

The 300-Ω TV ribbon can be purchased at many outlets including RadioShack and local hardware stores. Once again, if you want to use heavier-duty feed line, do so. The only proviso is that you may then have to trim the feeder length to be within tuning range of the Squirt's antenna tuner.

The End Insulators

I used 1/2×1 1/2-inch pieces of 1/16-inch PC board for the Squirt's two end insulators. As with everything else with the Squirt, these dimensions are not sacred; tailor them as you wish. If you use PC board for the end insulators, you have to remove the copper foil. This is easy to do once you've gotten the knack. Practice on some scraps before tackling the final product. The easiest way to remove the foil without etching it is to peel it off using a sharp hobby knife and needle-nose pliers. Carefully lift an edge of the foil at a corner of the board, grasp the foil with the pliers and slowly peel it off. You should become an expert at this in 10 or 15 minutes. Drill 1/8-inch-diameter holes at each end of each insulator for the element wires and tie-downs.

Tuner Feed-Line Connector

The tuner end of the feed line is terminated in a special connector. Because the TV-ribbon conductors aren't strong, they'll eventually suffer wear and tear.

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Trade-Offs

One of the unfortunate consequences of shrinking an antenna's size is that its electrical efficiency is reduced as well. A full-size dipole is resonant with a feedpoint impedance that matches common low-impedance coax quite well. This means that most transmitter power reaches the antenna minus only 1 dB or so feed-line loss. However, when the antenna is shortened, it is no longer resonant. A *NEC-4* model for the Squirt shows that its center impedance on 80 meters is only about 10 Ω resistive, but also about 1 kΩ capacitive. This is a horrendous mismatch to 50-Ω cable, and feed-line loss increases dramatically with high SWR. The Squirt uses 300-Ω TV ribbon for the feed line with an inherently lower loss than coax. This loss is much less than if coax were used, but it's still appreciable. Calculated loss with 300-Ω transmitting feed line is about 7.7 dB (loss figures are hard to come up with for receiving TV ribbon) so the feed line used doubtless has more than that.

Although this sounds discouraging, it's *not fatal*. You have to balance losing an S unit or so of signal against not operating at all! Consider that the Squirt, even with its reduced efficiency, is still better than most mobile antennas on 40 and 80 meters. So for local communication (a low-dipole's forte), using PSK31 and the Squirt is quite practical.

If you don't already have an antenna, the Squirt's a good choice to get your feet wet when using PSK 31. Once you get hooked, you'll probably want a better antenna. If you have the room, put up a full-size dipole; you'll see the improvement right away. If you can't do that, use a lower-loss feeder on the Squirt, such as good-quality open wire.—*Joe Everhart, N2CX*

Figure 4—The pad side of the home-made feed-line-to-tuner connector.

Figure 5—Here the feed-line-to-tuner connector is shown attached to the binding posts of the Squirt antenna tuner.

This connector provides needed mechanical strength and a means of easily attaching the feed line to the tuner. In addition to some PC-board material, you'll need four or five inches of #18 to #12 solid, bare wire. Refer to Figure 3 and the accompanying photographs in Figures 4 and 5 for the following steps.

Take a $1\frac{1}{8}\times1\frac{3}{4}$-inch piece of single-sided PC board and score the foil about $\frac{1}{2}$inch from one end; remove the $1\frac{1}{4}$-inch piece of foil. Now score the remaining foil so you can remove a $\frac{1}{8}$-inch-wide strip at the center of the board, leaving two rectangular pads as shown in Figures 3B and 4. Drill two $\frac{1}{16}$-inch-diameter holes in the copper pads spacing the holes about $\frac{3}{4}$-inch apart. Drill two $\frac{3}{8}$-inch holes at the connector midline about $\frac{5}{8}$-inch apart, center to center, to pass the feed line and secure it.

Cut two pieces of #18 to #12 wire each about three inches long. Pass one wire through one of the $\frac{1}{16}$-inch holes in the connector board and bend over about $\frac{1}{4}$-inch of wire on the nonfoil side. Solder the wire to the pad on the opposite side and cut the wire so that about oneinch of it extends beyond the connector. Repeat this procedure with the second wire. Next, strip about two inches of webbing from between the feed-line conductors and loop the feed line through the two $\frac{3}{8}$-inch holes so that the free ends of the two conductors are on the copper-pad side. Strip each lead and solder each one to a pad. You now have a solid TV-ribbon connector that mates with the binding-post connections found on many antenna tuners. Figure 6 shows the connector mated with a Squirt tuner.

Center Insulator

Strip all the foil from this 3-inch-square piece of board. Use Figure 3A as a guide for the hole locations. The top support hole and the six wire-element holes are $\frac{1}{8}$-inch in diameter; space the wire-element holes $\frac{1}{4}$-inch apart. The feed-line-attachment holes are $\frac{3}{8}$-inch diameter spaced $\frac{1}{2}$-inch apart, center to center; the two holes alongside the feed-line-attachment holes are $\frac{1}{16}$-inch diameter. These $\frac{1}{16}$-inch holes accept a plastic tie to secure the feed line. I trimmed the insulator shown in Figure 2 from its original 3-inch-square shape to be more esthetic. Your artistic sense may dictate a different pattern.

Bevel all hole edges to minimize wire and feeder-insulation abrasion by the glass-epoxy material. You can do this by running a knife around each hole to remove any sharp edges.

Putting It All Together

The Squirt is simple to assemble. Once all the pieces have been fabricated, it should take no more than an hour or two to complete assembly. Begin with the center insulator. Cut each of the two element wires to a length of about 34feet. Feed the end of one wire through the center insulator's outer hole on one side, then loop it back and twist around itself outside the insulator to secure it. Now loop it through the other two holes so that the inner end won't move from normal movement of the wire outside the insulator. Repeat the process for the other insulator/wire attachment. Separate several inches of the TV-ribbon feed-line conductors from the webbing; leave the insulation intact except for stripping about $\frac{1}{2}$inc h from the end of each wire. Pass the TV ribbon through both $\frac{3}{8}$-inch holes. Strip a $\frac{1}{2}$-inch length of insulation from each dipole element, then twist each feeder wire and element lead together and solder the joints. It might be prudent also to protect the joint with some non-contaminating RTV or other sealant. Finally, loop a nylon tie through the holes alongside the feeder and tighten the tie to hold the feeder securely. A close-up of the assembled center insulator is shown in Figure 6.

Attach the end insulators to the free ends of the dipole wires by passing the wires through the insulator holes and twisting the wire ends several times to secure them.

So that the antenna/feed-line system can be tuned with the Squirt tuner, the 300-Ω feed line needs to be about 45 feet long. If you use a different tuner, you may have to make the feed line longer or shorter to be within that tuner's impedance-adjustment range.

Tuner Assembly

This tuner (see Figures 7 and 8) is about as simple as you can get. It's a basic series-tuned resonant circuit link-coupled to a coaxial feed line. At C1, I use a 20 to 200-pF mica compression trimmer acquired at a hamfest (you *do* buy parts at hamfests, don't you?), although almost any small variable capaci-

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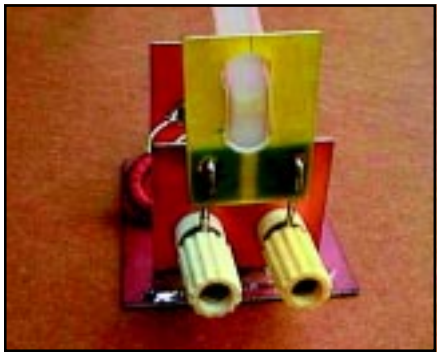

Figure 6—View of an assembled center insulator fashioned from a 3×3-inch piece of PC board from which all the foil has been removed.

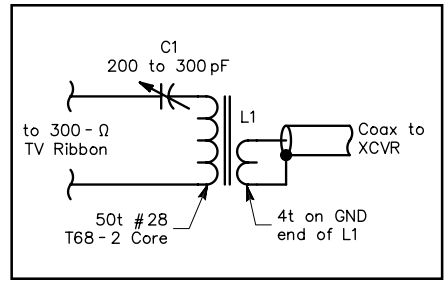

Figure 7—Schematic of the Squirt antenna tuner. See the accompanying Parts List.

Figure 8—This Squirt tuner prototype uses a 2×3-inch piece of PC board for the base plate, two 1½ × 1½-inch pieces for end plates and a ½-inch square piece as a tie point for the toroid and tuning capacitor.

tor of this value should serve. The inductor, L1, consists of 50 turns of enameled wire wound on a T68-2 iron-core toroidal form. An air-wound coil would do as well, although it would be physically much larger. Figure 8 shows the tuner built on an open chassis made of PC board. My prototype uses several PC-board scraps: a 2×3-inch piece for the base plate, two 1½×1½-inch pieces for each end plate (refer to Figure 3). A ½-inch square piece of PC board (visible just beneath the capacitor in Figure 8) is glued to the base plate to serve as an insulated tie point for the connection between the toroid (L1) and tuning capacitor (C1). The tuner end plates are soldered to the base plate to hold a pair of five-way binding posts and a BNC connector at opposite ends. L1 and C1 float above electrical ground, connected to the TV ribbon. One end of L1's secondary (or link) is grounded at the base plate and the coax-cable shield. The hot end of L1's secondary winding is soldered to the coax-connector's center conductor.

Tuner Testing

C1 tunes sharply, so it's a good idea to check just how it tunes before you attach the tuner to an antenna. You can simulate the antenna by connecting a 10-Ω resistor across the binding posts. If you use an antenna analyzer as the signal source, a ¼-W resistor such as the RadioShack 271-1301 is suitable. But if you use your QRP transmitter, you need a total resistance of 8 to 10 Ω that will dissipate your QRP rig's output, assuming here it's 5 W or less. Four RadioShack 271-151 resistors (two series-connected pairs of two parallel-connected resistors) provide a satisfactory load if you don't transmit for extended periods. Or, you can make up your own resistor arrangement to deliver the proper load. Adjust C1 with an insulated tuning tool to achieve an SWR below 1.5:1.

Once the tuner operation is verified using the dummy antenna, it's ready to connect to the Squirt. Tuning there will be similarly sharp, and a 2:1 SWR bandwidth of about 40 kHz or so can be expected as normal.

A Multiband Bonus

Although the Squirt was conceived with 80-meter operation in mind, it can double as a multiband antenna as well. The simple Squirt tuner is designed to match the antenna only on 80 meters. However, a good general-purpose balanced tuner such as an old Johnson Matchbox or one of the currently popular Z-match tuners (such as an Emtech ZM-2) will give good results with the Squirt on any HF band. The Squirt prototype was recently pressed into service at N2CX on 80, 40, 30, 20 and 15 meters for several months. It worked equally as well as a similar antenna fed with ladder line. Although no extensive comparative tests were done, the Squirt has delivered QRP CW contacts from coast to coast on 40, 20 and 15 meters and covers the East Coast during evening hours on 80meters.

Build one! I'm sure you'll have fun building and using the Squirt!

Parts List

Squirt Antenna

Numbers in parentheses refer to vendors presented at the end of the list.

1—3×3-inch piece of 1/16-inch-thick glass-epoxy PC board for the center insulator (1)
2—½×1½-inch pieces of PC board for the end insulators (1)
1—1⅛×1¾-inch piece of PC board for the feed-line connector (1)
2—34-foot lengths of #20 (or larger) insulated hookup wire (2)
1—6-inch length of #16 (or larger bare) copper wire; scrounge scraps from your local electrician.
1—45-foot length of 300-Ω TV ribbon line (2)

Squirt Tuner

1—2×3-inch piece of PC board for base plate (1)
2—1½-inch-square pieces of PC board for end plates (1)
1—½-inch-square piece of PC board for the tie point (1)
1—200- to 300-pF (maximum) mica compression trimmer (3)
1—T68-2 toroid core (3)
2—Five-way binding posts (2)
1—55-inch length of #26 or 28 enameled wire (2 and 3)

Note: You can use 3/16-inch-thick clear Plexiglas for the Squirt's end and center insulators. Commonly used as a replacement for window glass, Plexiglas scraps can be obtained at low cost from hardware stores that repair windows.

Vendors

1. HSC Electronic Supply, 3500 Ryder St, Santa Clara, CA 95051; tel 408-732-1573, **www.halted.com**
2. Local RadioShack outlets or **www.RadioShack.com**
3. Dan's Small Parts and Kits, Box 3634, Missoula, MT 59806-3634; tel 406-258-2782; **www.fix.net/~jparker/dans.html**

Notes

[1]Dave Benson, NN1G, and George Heron, N2APB, "The Warbler—A Simple PSK31 Transceiver for 80 Meters," *QST*, Mar 2001, pp 37-41.

[2]R. Dean Straw, N6BV, *The ARRL Antenna Book* (Newington: ARRL, 1997, 18th ed), pp 5-9 to 5-11.

[3]**www.alphalink.com.au/~parkerp/nodec97.htm**; **www.home.global.co.za/~tdamatta/loops.html**

[4]**home.earthlink.net/~mwattcpa/antennas.html**

[5]Full-size templates are contained in SQUIRT.ZIP available from **www.arrl.org/files/qst-binaries/**.

You can contact Joe Everhart, N2CX, at 214 New Jersey Rd, Brooklawn, NJ 08030; **n2cx@arrl.net**.

Photos by the author

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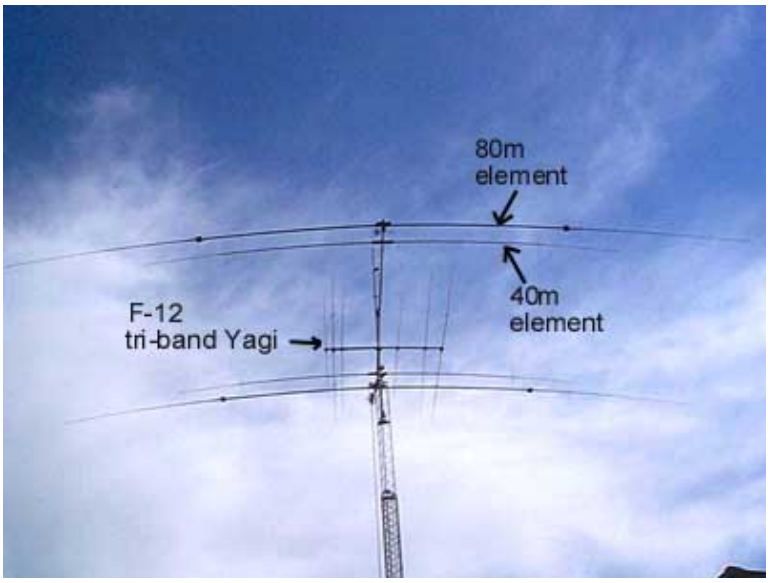
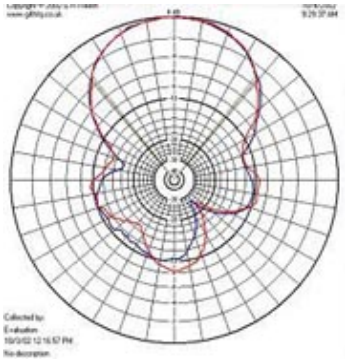


80m & 40m DUAL BAND HOMEBREW 2 EL SHORT BOOM YAGI

Follow the links at the left to see various pages of design and construction details. (Click to see a full size photo of the Yagi)

The VE6WZ QTH is a small city lot, so achieving gain on 80m has been difficult. A vertical 4 sq. wire array was considered but would have been a compromise design because of the crank-up tower and limited space. The surrounding ground clutter and variable terrain would have made vertical performance questionable. Good DX results have been experienced using the Force-12, EF-180B, 80m rotatable dipole at 100' so it was decided to design a 2 el 80m yagi around similar elements. (F-12 markets such yagis). The VE6WZ yagi design uses high Q "mostly air core" loading coils instead of the linear loading on the 68' elements. Because of dimensional constraints at the VE6WZ city QTH, a 28' short boom reflector design was built with the 2 el 40m yagi sharing the same boom.



Above: Field measured azimuth plot of the Yagi on 3.8 and 3.793 MHz.

80m 2 el Yagi specifications

Max F/B:	20 dB peak (10 dB within SWR bandwidth)
Max gain:	8.95 dBi (at 100', 33 deg ele. angle)
2:1 SWR bandwidth:	26 kHz (with switching 55 kHz) 3810-3760, 3550-3500.
Feedpoint impedance:	15-20 ohms (hairpin match)
Longest element:	66'
Boom :	28' (2-1/2" X .125) Trussed
Weight of 80m only:	170 lbs
Weight total with 40m elements:	210 lbs
***Estimated* windload:**	13 sq. feet

Working DX is not proof of good performance, however in the first 7 months of the 2002-3 winter season from Sept.1 to Mar.31 at VE6WZ 112 DXCC countries were worked on 80m, and 187 were worked on 40m.

Listen to the directivity of the Yagi on 80m: In this QSO with Jose, F5JD at first the Yagi was pointed at 90 deg. with very poor copy. Part way through the recording the beam is rotated to direct path at about 40 deg. and the copy is Q-5. This IS NOT a F/B test...the beam is only rotated about 50 deg. Click here and your sound player should play this .mp3 file: F5JD_80m-03-10-12.mp3

More sound recordings can be heard on the Sound File Page.

Compromised performance issues (80m):

This Yagi is almost ½ full size on 80m and is a great DX performer. It has about 1 dB less gain but shows *better* F/B than a design using full size elements (see here). The small size makes it feasible to install in many situations where a full size monster would not be possible.....however some performance and operational issues arise:

1.) This would NOT be a great contest antenna because of the *very narrow bandwidth* and requirement to use switching to move around the band. The bandwidth could be expanded by retuning the reflector to sacrifice some gain, but some switching would still be required.

2.) The large coils are susceptible to inductance change when covered with snow, ice or frost. In severe cases this can render the Yagi unusable.

3.) Because of the narrow bandwidth and tight coupling of the elements the SWR will "swing" during windy conditions as the elements move.

4.) The narrow bandwidth and close coupling of the elements makes tuning critical. Great care and effort must be taken to adjust the elements for best performance. On the tower, field testing will almost certainly be required to get the tuning right. **WARNING...This is not a "plug-and-play" design.** If you build this design be prepared to spent much time tuning the elements for optimum performance.

(see here for comparison plots to a full size design)

WHY INDUCTOR LOADING ??

VE6WZ displays one of the 80m loading coils.

The Cushcraft "shorty-forty" experience.....

The Cushcraft XM-240, 2 el 40m yagi has been used at VE6WZ for 3 years. It has 43' inductor loaded elements mounted on a 22' boom. The loading coils are 68 turns of 12 ga. wire close-wound on a small diameter (3/4 inch) fiberglass form. Other sources have indicated a coil loss of around 3 ohms, and software calculations indicate a loss of up to 8 ohms !! (see photo of the Cushcraft coil). The coils are placed about 61 % out from the center along the element half length. The split element is fed directly with 50 ohm coax through a 1:1 balun. In spite of these poorly designed lossy coils, this yagi is a proven performer. The thesis is that an 80m design using well designed high Q coils should perform well.

The Force 12 linear-loaded 80m dipole experience.....

The F-12 80m dipole (EF-180B) is a 68' long, linear loaded element and has performed well at VE6WZ for two years. The linear load wires extend from the mid-point of the half element inward toward the element center. The wires are aluminum clad 12 gauge steel wire. The split element is fed with 50 ohm coax in combination with a "helical hairpin" to match the 18 ohm feedpoint. MININEC was used to model an 80m dipole using 68' elements and lumped inductance (coils) at 50 % from center. The feedpoint impedance is calculated at around 32 ohms. This is almost double the 18 ohms realized when using linear loading where the wires come in toward the feedpoint. Others have explained that this low impedance is because the "effective" load point is moved inward toward the center due to the loading wires running inward along the element. A low feedpoint impedance will tend to decrease efficiency. Tests on the tower support the 32-ohm feedpoint using low loss (high Q) coils. The 2el yagi design using 18 ohm linear load elements would end up with around a 10 ohm feedpoint !!!! due to mutual coupling. This could greatly degrade efficiency. Modeling the 2 el inductor loaded design yields a reasonable 16-20 ohm feedpoint. The total wire length used in the coils is much less than that needed for linear loading, and the ¼ " copper tubing used for the high Q coils has less resistance than the 12 gauge AL clad wire. The element connection points can introduce further resistive losses and these are minimized at the coil-element connections by implementing electrical redundancy.

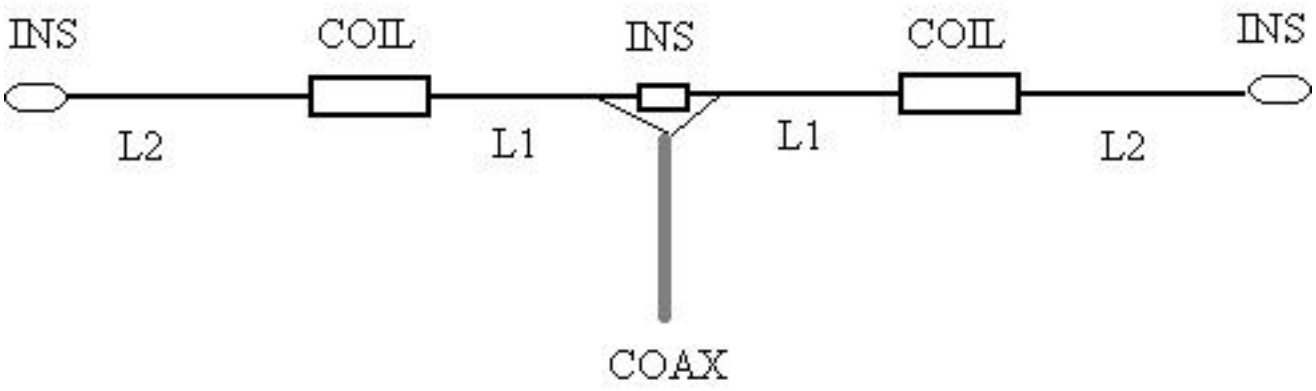
NEXT.....Design and modeling

Below...the Yagi at 100' at sunrise.

13883

A SHORT DIPOLE FOR 80m

4S7NR



The antenna above has been described by Nadisha, 4S7NR and may be of interest to anyone wishing to get on 80M (3.5MHz) that have limited space available.

L1 is 12 feet. L2 also is 12 feet and the overall length is 48 feet.

The two loading coils are described as 67.83uH and can consist of 104 turns of insulated wire, wound over 3.5 inches. The coil diameter is not stated however. Maybe it will be a case for experimentation here.

I have a file, a dipole for 20M (Click to view) which also is a loaded dipole and works well.

It would be interesting to try loaded dipoles for other bands. I will have a try at one for 40M soon.

if you have any ideas on such dipoles or experiences with them, let me know.

(Type a title for your page here)

Back to the first page

Harry's Homebrew Homepage

Start Pages	Projects	Reference

Welcome to the "new look" homepages. Above are your subject areas, below left is the file menu and below right you will see the project file selected from the left menu. NEW Latest addition is DIY IF Cans for Valves

Start menu

Sommarstuga till salu
Start Page
Jag söker jobb UPDATED
Messageboard
Project kits

Links/Suppliers
Publish/Copying
Homepage info
My wife and me
Faq + Fasq
E-Mail me ?? OLD
What's new UPDATED

Harry Lythall SM0VPO/G4VVJ

Updated 23rd July 2004

Technical questions? Visit the Messageboard.

Welcome to Harry's Homebrew Homepages; a resource for radio, hamradio and electronics in general. I hope you have fun with my homepages, regards from Harry - SM0VPO, (Lunda), Sweden.

Sites Summary

Homebrew Projects are located at:

http://w1.859.telia.com/~u85920178/	Unlimited bandwidth, Full site, No advertising, SWE

Visitors Circuits are located at:

http://hem.spray.se/lythall/	Selected visitors submitted circuits - SWE

A small PDF library is located at:

http://hem.passagen.se/sm0vpo/	Unlimited bandwidth, Full site, SWE

= Photograph illustrations included with the project.
UPDATED = I have recently updated this project (Obvious, really!!!)
NEW = Recently added (no timeframe - I'm quite informal)
OLD = Not new (but I want it to be noticed by someone)
= The picture file is missing - please let me know.

Here you can e-mail Harry, but if it is a technical question then **please** be so kind and use the MessageBoard. Thank you.

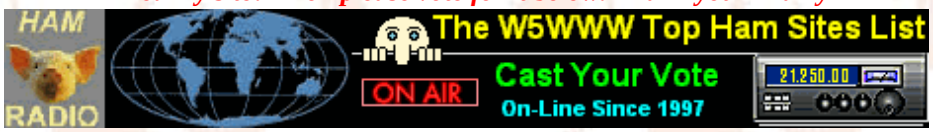
AMATEUR RADIO BOATANCHORS WEB RING

Join BA Ring	This BOATANCHORS Webring site is owned by Harry Lythall. Want to join the BOATANCHORS Webring?	Next BA Site

[Prev] [Next] [Skip Next] [Random] [Next 5] [List Sites]

Visitors since 01st Jan 2003: 261291 (186,000 during 2003).
FastCounter by bCentral

Liked my site? Then please vote for it below. Thank you - Harry



W5WWW's Top Ham Sites List

Boycott Internet Spam!

"Taylor Space Miser 80m antenna"

Roger Harrison (after James Taylor W2OZH)

Published: "Australian Electronics Monthly" magazine, September 1987

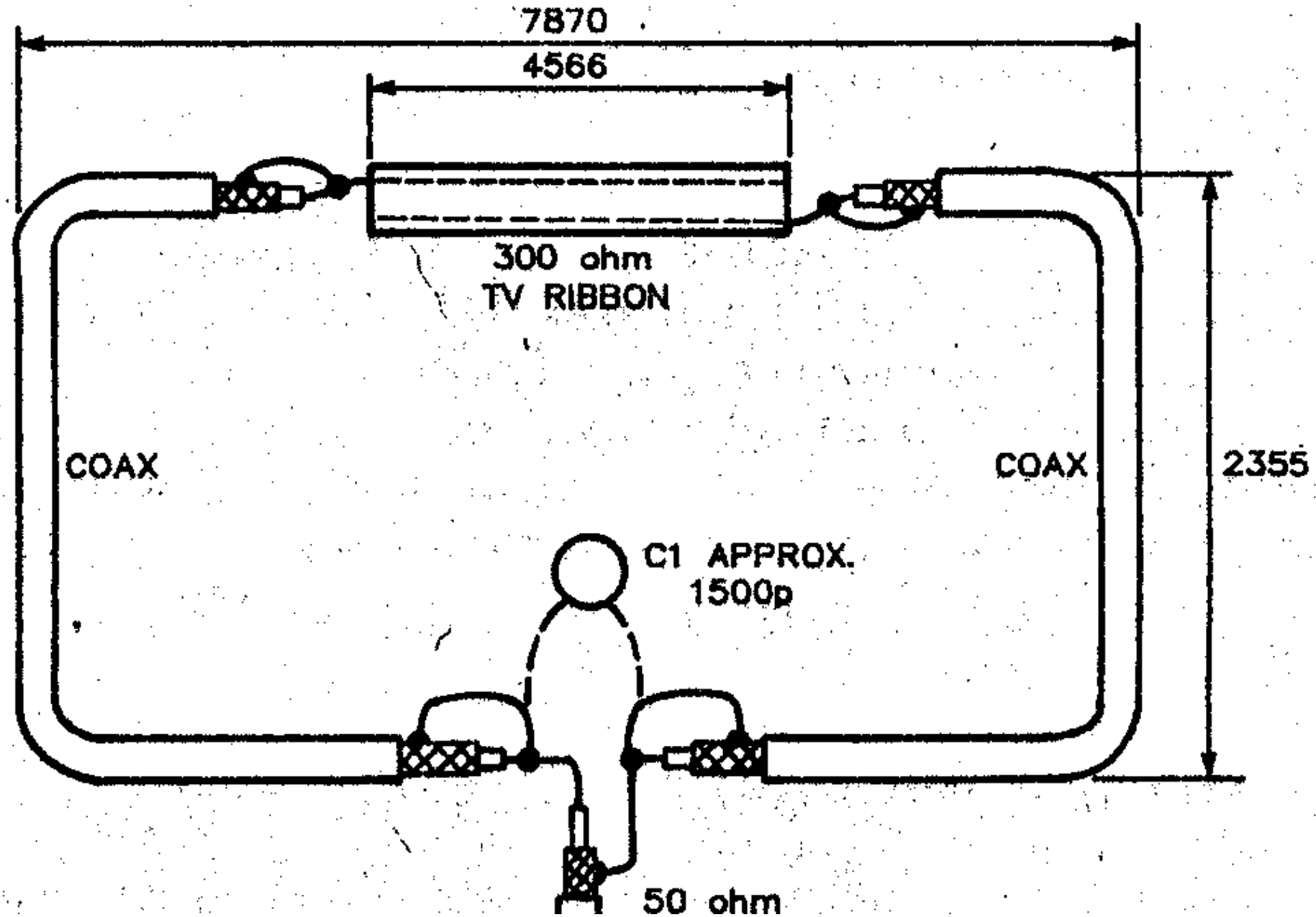


Figure 1. The general form and dimensions of the antenna. Now you can see why it's called the "space miser"!

PLEASE NOTE:
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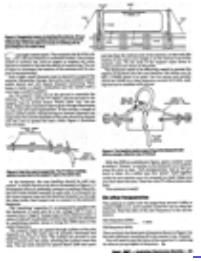
Among other ways to contact Roger Harrison, you can also do so through the author this web page, at prellis@pcug.org.au]

176kB



180kB

(January 2004) This antenna has been given new life in Canberra's Amateur Radio circles through the efforts of Tony Bennett VK1TB. Please look back for Space Miser 80m dimensions that he is developing for a "town-house roof-space".

"Here are the results with my antenna mounted in the ceiling.

Taylor Loop 80M antenna RG-213 or other 10mm co-ax.

8.120M f=3.854MHz
9.370M f=3.580MHz
9.640M f=3.531MHz

300R TV ribbon 4.65M

Bandwith at SWR=1.5: 3.59 - 3.53 for 9.370M co-ax."

Tony's loop is tilted because it follows the pitched roof of his town-house. Note: There is no aluminium-foil lining in his roof.

I also repeat here the text of an email message I received from Roger Harrison.

Hi Peter,

Thank your for the email.

I am happy for the Space Misrer antenna and my article to be given a "new lease of life."

However, the proprietaries should be observed.

The Copyright Act 1968 governs use of the article. Any "work" printed, published or produced by some means is automatically copyright. No registration is required. Copyright can be sold or licensed by the creator of a work.

Although Australian Electronics Monthly is no longer published, I have retained copyright in all my "works" published in AEM (and, incidentally, a lot of other material). Similarly, I have retained copyright in all articles authored by me and published in Electronics Today.

Hence, I hold the copyright to the Space Miser antenna article published in September 1987 AEM.

However, I am happy to give you permission to reproduce the article on your website, as-published (ie. scanned pages reproduced as images), with the proviso that you add this note:

Copyright 1987 Roger Harrison VK2ZRH. Reproduced with permission. This work is not to be mirrored or otherwise reproduced by electronic means without permission of the author.

With that, the proprietaries are observed. Good luck experimenting with the antenna. I built one and mounted it on the outside back wall of our (then) office in Balmain - which overlooked a rear balcony. Nothing flash, but it did the job remarkably well.

73, Roger Harrison VK2ZRH

"Taylor" Space Miser 80m antenna

Roger Harrison (after James Taylor W2OZH)

Published: "Australian Electronics Monthly" magazine, September 1987

Love My Loop!

A Small Loop Antenna
for Forty Meters

By Ben Smith, W4KSY

I promised Jack Stone, Publisher of ***antenneX***, if the short loop worked as well as my phased verticals on forty meters, I would write an article. I didn't promise a GOOD article because I am no writer, but it is a GOOD antenna. It does perform as well as my phased verticals. I am satisfied the effort in building this test model was worthwhile. Here is an attempt to tell you about my experiences, including mistakes. It was challenging and fun. Perhaps this will spur you into taking on the endeavor of building a loop. Time flew, and it certainly renewed my interest in Ham Radio. There is nothing like a new antenna for a conversation piece. You get extra bragging rights when you build it yourself. (My friends hate me!) If you do bite the bullet and build one, it is hoped you will save frustration, time, and money. If you don't build it you will have a good time thinking about what a nut this guy must be! So, read on!

THE LOOP BOOK

I clicked on an ***antenneX*** link and read some of the free stuff. I wanted to know more about "shortened loops," and bought ***The Loop Book*** found in the Shopping Shack on this web site. After reading, or rather, "Screening," the book, the decision was made to build a short loop. I read about the "Special Capacitor" requirement—the need for welded construction, and of course, the capacitor has to be controlled remotely. I did a little research and found that suitable vacuum capacitors cost a lot of money. I wasn't going to invest any more than I had to in order to see how well one of these Shortened Loops worked. A stepper motor and control would cost a hundred bucks plus the cost of the capacitor. ("A steel shaft in the capacitor inserts too much loss!!!!") Hmmmm!

CAPACITOR CAPER

I am a creature of habit. I only work in a small portion of each band most of the time. For an eXperimental antenna, I could use a trimmer capacitor. I don't need to go from one end of the band to the other. Servo, and/or stepper motors and remote controls are not my cup of tea. The capacitor has to withstand high voltage and current, be efficient, and adjustable. I ended up building a capacitor out of a one inch piece of copper tubing sliding over a piece of half inch copper tubing that had an insulator separating the tube into two pieces. The separator is made of acrylic tubing inserted a little off center. The spacing for the two tubes was facilitated by making two acrylic washers, or bushings, to fit in the end of the one inch pipe with half inch holes so it would slide over the smaller tubing. ***Voila!*** A concentric or ***trombone*** capacitor!

I had a drill press, the plastics store is three miles away, Home Depot is four miles away, and I had a few bucks in my pocket. I bought the tubing (one inch copper tubing costs a little over a dollar a foot), took my treasure home and the fun began. But, the hole saws on hand were not suitable. Three hole saws were purchased, one was a mistake, (a little over eleven dollars wasted!) and the air soon was filled with the aroma of burning plastic—cough, cough! Acrylic melts! Slow and easy are the watchwords. Now how does one secure three eighths of an inch acrylic washer in copper tubing? Drill some holes through the edge of the tubing and into the washer which is epoxied in place and screw in some little nylon screws, also imbedded with a little epoxy. One must remember to have everything working before you glue and screw.

I found some acrylic rod that fits into the smaller copper tubing. I pushed one piece of the half inch tubing over the rod and then pushed in the other piece, leaving about a half inch of space in the middle. I used some epoxy on the rod before assembling the "inner plate." Then, for good measure, a center punch was used to peen the tubing in a few places to make sure that it was going to stay together. Glue is good, but mechanical means is better. I found a formula for capacitance somewhere and estimated I would have a fifty pF capacitor. It turned out to be forty-four picofarads. Which means that only twenty-two of it is useable, because it is used as a split stator capacitor. No wipers, no contact resistance. Efficient air dielectric with quarter inch spacing. ***Ain't dat sumpin!***

GREAT MINDS?

A little shy of C for my design. ***My design?*** I took the forty foot job (forty feet in circumference, a little over sixteen feet in diameter) from ***The Loop Book***. I thought I had created something when I built that capacitor. Well, don't you know, the very next issue of ***antenneX*** had a trombone capacitor in it!!!! However, I had an air dielectric, which is nicer, neater and more efficient than plastic tape—but, on the other hand, the guy didn't have to have a drill press and access to acrylic, nylon screws, epoxy, etc. If I remember properly my capacitor was about forty-five inches long. So you get approximately one pF an inch for one inch tubing over half inch tubing. You had better do your own numbers though if you are going to take on one of these things because I am not an engineer. Enough of the capacitor.

BUILD THE LOOP

So get out the butane torch, flux, emery cloth, tubing cutter, shine and clean tools, for three quarter inch tubing—eight pieces of three quarter inch copper tubing (I used thin wall...another mistake.) I Should have used the thick wall stuff and one inch tubing. This thing looks and feels flimsy when you try to pick it up and get it in the air. Supports could be made from wood and acrylic insulators, but why go to the trouble if it will hang together long enough for testing??? No more mechanical work than necessary is my motto. Besides the elbows I had to get a couple of tees, some reducers from three quarter to half inch

tubing. Why the reducers? To mount the capacitor on top and, parallel to the topmost member of the loop. Also, the top member of the three quarter inch tubing has to be broken and spaced in the center with another piece of acrylic rod. I had some plastic milk crates which served as supports for the loop while it was under construction and also testing after it was built.

CLICK TO LOAD PICTURE OF LOOP ON CRATES

PLUMBING SPECIAL

Now, here was the easy part, a plumbing special. Be sure to get everything clean and shiny. Where the tubing goes into the elbow, do a good job on the outside of the tubing and the inside of the elbow. Apply a little, but uniform coating of flux on both members. Apply heat to the elbow. I was surprised to see how well it sucked up the solder. I had done a little plumbing before and it was messy. But by heating up the outside member it sure pulls in the solder and it makes a neat job. A judicious wipe (careful) while the solder is molten will get rid of any pile-up.

Holy Buckets! Here is a forty foot octagonal of copper pipe with a capacitor in the middle. It looks real different out in the backyard. My cat came out to supervise on good days. He would sit up on a milk crate and sneer and make snide meows.

FEEDING

Now how do we feed this thing? ***The Loop Book*** suggested three ways. Well, before we match an antenna, we have to make it resonant at the desired frequency. I have built many antennas, dipoles, coaxial, bazooka, verticals, quads, rope Yagis—so, I know all about this stuff. I got out my homebrew noise generator, took about ten turns around the bottom of the loop. I connected it, then took another ten turns and coupled it to my coaxial feedline. Then, went in the house and turned on the receiver to find the resonant frequency of the loop. I was measuring the resonant frequency of the coupling coils, the feedline, the response of the receiver, ***everything except the loop.***

Okay another tack! Got out my fancy DDS VFO with digital read out and my trusty Fluke digital MM with RF probe. I just about wore out the knobs on the VFO. I was getting peaks, but none of them were where they should fall. One consistent QRG (I am a CW operator, ahem!) was 6960. Moving the capacitor didn't change the frequency, but the peak was broad. Maybe the loop was too long. I took out two members from opposite sides and sawed out a foot, then two feet. No change.

Maybe, I need a dip meter. Couldn't find one locally. Build it...? Yeah, bought some banana plugs for the coils, searched and found suitable capacitor, knob, dial, box. Then while surfing The Internet one evening, I ran across the homepage of a Ham Distributor out in the Midwest. Sent him an E-Mail and whatdya know, they had one. I Bought a dip meter.

While awaiting the delivery of the dip meter, the Publisher of ***antenneX*** sent me an E-Mail telling me that I had received a compliment on my Over/Under Quad article published in the November 1997 issue. I sent him a reply of thanks and told him the blankety-blank loop was driving me up the wall because I

could not find its resonant frequency. A few hours later, Jack sent me an E-Mail saying that the old maestro of loops, K5CNF, Richard Morrow; said I wouldn't find it by the methods I was using. I would need to put in a matching system and go for lowest VSWR. Now that really blew my mind!

So the Dip Meter arrived. Hot Dawg! I decided to use the tapped capacitor method of feed as outlined in ***The Loop Book.*** I went down to the hardware store, bought some quarter inch copper tubing, wrapped it in plastic tape, flattened one end, drilled a hole in it, then drilled a hole in the loop, secured that end with a self tapper. After I found the proper place for the tap, the tubing could be soldered to the loop—how clever! I taped the quarter inch stuff to the loop and mounted a coaxial connector on the loop by soldering on a strap. I connected the other end of the quarter inch tubing to the center of the connector with a piece of heavy flexible wire inserted in the tubing, and then crimped and soldered.

MAKING CONNECTION

The tubing came in a twenty-one foot length. So, how far out from the connector do I go? I figured, it would be better to start long than to be short and have to splice. I used the whole thing! After drilling numerous holes, cutting, flattening and boring holes in the tubing, I was within two feet of the connector and no pay-dirt. Once or twice I would see a small drop in VSWR. But moving the tap either way, or spacing the matching section away or closer to the loop did not make the desired effect!

I bought an MFJ VSWR Analyzer—two hundred bucks !—but, great piece of equipment. Now I had the whole mess of test equipment in one box with a VSWR Meter and a Resistance Meter side by side, a signal generator with a digital readout. Alas, it told me nothing new.

GAMMA MATCH

The next trial was going to be by the pick-up loop method. One loop shown in ***The Loop Book*** had a pick-up loop one-fifth the size of the main loop. Okay! I need an eight foot loop to place inside my forty-footer. I had the tubing cut, elbows in place and ready to solder when a light came on. How about a ***gamma match?!***

Wow! Found an old ARC-5 Transmitter Capacitor. It was nicely made with good wide spacing. I measured the capacitance with my new capacitance meter. It was 144 pF max. I mounted it on a piece of acrylic sheet, fastened the sheet to the loop with a couple of straps, recovered the remains of the quarter inch tubing and went to work. BY GOLLY it worked. A few taps up and down, pulling the small tubing away from the larger one, shoving it back and forth, and adjusted for resonance at the desired frequency with my highly efficient, and beautiful homebrew capacitor. Inside of five minutes after installation, I had an honest-to-heavens one-to-one match, with 50 ohms indicating on my shiny new analyzer at 7020 kHz!

I pulled off the analyzer, connected the feedline, ran in the house, plugged the analyzer into the other end of the feedline and whoopy-de-do!—one to one and 50 ohms. A little "smidgin" over one to one, but it wasn't to the first calibration point. We won't quibble over One point, Oh, One to One!

SIGNALS GALORE!

Next steps: plug in to the coaxial switch. The switch facilitates connecting to an eighty meter dipole and a phased vertical array for forty. Then connect the coaxial switch to my Icom 765. A quick check to where I set the lowest VSWR-7020 KHZ. Sure enough the ICOM'S SWR meter agreed. I hear signals. Lots of signals. Spent the next couple of hours making comparisons between received signal strength on the loop and my phased verticals—aiming the verticals at the received station. The loop was two S units down most of the time relative to the verticals. On occasion the loop was as good. Most of the time the verticals were better. ***But,*** the loop was only supported by plastic milk crates, laying down horizontally to the earth and just one foot above it!

While making a VSWR check at reduced power (eight watts) I sent three Vee's de W4KSY. I got a call from K1LGQ. Dennis, in Nashua, New Hampshire; gave me a 599 report before he knew I was on a loop and aching for a decent report. I must admit that I went to full 100 watts when I responded to his call. Never-the-less, I was elated. I made several more contacts and received good reports compared with the received signals. W9OVY, Wally near Charleston, S. C., was most helpful in assisting with tests very early in the game.

HAUL TO VERTICAL POSITION

Somewhere along this adventure I had decided the tuning capacitor was coupling to the top member of the loop, since it was only a few parallel inches away. I cut it out! Later I replaced it with an acrylic rod. This will be a great insulator to use to support the antenna! Ho!

Next day I was out in the backyard with slingshot, line, pulley, etc., etc. Got a line , with pulley attached, over a limb and hauled the loop up into a vertical position. No mean feat, single-handed. That thing sure is wobbly with no extra support. (Note to me: NEXT TIME: USE ONE INCH TUBING FOR A FORTY FOOT LOOP, or heaven-forbid, use insulated wooden supports.) Got out my trusty Analyzer, retuned to one-to-one quickly, then went inside and had a ball. Numerous contacts were made with good reports.

EARLY COMPARISONS

Summary: In most cases the loop was as good as, or superior to the phased verticals at distances out to about 800 to 1000 miles. Then the verticals seemed to take over. There were times when, even with long haul DX, the loop was better by an S Unit or two. Remember, I was aiming my phased verticals. The pattern is broad, but worth about an S Unit for locals and sometimes three S Units for DX. The loop was fixed north and south. For the past few days November 18th to November 22 there has been deep QSB,

even in daylight hours. This afternoon a South African Station at 7010 was S5 on my phased verticals and S5 on the loop. The QSB seemed slightly more prevalent on the loop then on the verticals. Background noise was about the same. Some QRM from US Stations west of me, and in the null of the phased array, was less on the array and more on the loop, of course. Figures! I hang around 7020 kHz with keyboard, or keyer, and LOOP. Demonstrations gladly given.

ANOTHER LOOP PLANNED

OKAY! This is Mark One! It is a good antenna. The performance is better than I expected. Therefore: MARK TWO is in the works. I have a new acquisition. I bit the bullet and bought a 5 to 5000 pF @ 15000 Volts, Vacuum Capacitor; with gearing, motor, limit switches and positioning pot. That monster weighs eight pounds, is almost a foot in length and five inches in diameter. The capacitor lists for $900.00! I got the whole thing Surplus for $187.00. Did I hear someone say something about anchoring boats? At ***antenneX,*** Richard, K5CNF, says to be careful with vacuum capacitors. Sometimes they are unable to handle the high current produced by the loop. According to ***The Loop Book***, I should be able to make a loop forty feet in circumference, sixteen feet in diameter work on 40, 80, and 160, with my vacuum capacitor, providing it will handle the current. ***Here I go, again!*** **-30-**

Send mail to webmaster@antennex.com with questions or comments about this web site.
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N3FJP World Class Station & Contest Loggers

GOOD STUFF

HOME

QSLWorks

DEDICATION

THE OP'S

WHY HAM RADIO - WHY NOT CB?

YOUR FIRST RIG

HF AMPLIFIER BASICS

GROUNDED GRID BASICS

POWER SUPPLY BASICS

WHAT NOT TO DO AROUND HV!

REFERENCE MATERIAL

WELCOME!

QSLWorks
Full Color Custom QSL Cards
(click here)

You have reached the original WB0NNI scratch built RF amplifier site. Many changes remain to be completed and will at some point in the future (probably in the dead of winter, when the beam is caked with ice and the space heater in the shack fails to break the frost off the rig) be completed. For now, however, you'll find a number of resources available for the homebrew design and construction of HF power amplifiers that employ glass triodes, tetrodes and pentodes (not that we're opposed to solid state amplifiers that have the look and and feel of a "George Forman Grill" - it's just that we're partial to equipment that looks like it might actually do something other than cook weiners at a backyard barbeque). With that said, we remain committed to our original mission of resurrecting the practical and cultural aspects of ham radio, including a glimmer of the "tongue-in-cheek" humor, fun and excitement that has, for so many years, attracted innovators of all ages to the hobby.

With the exception of our endorsement of N3FJP's neat, easy to use Windows logging software, and our recently

MORE GOOD STUFF

SELECTING A TUBE

RF INPUT CIRCUIT

RF OUTPUT (TANK) CIRCUIT

TRANSFORMER POWER CAPABILITIES

PARTS IS PARTS

MOSLEY CL-33 MANUAL

ROLLING YOUR OWN - TRANSFORMERS AND CHOKES

POWER SUPPLY PROJECTS

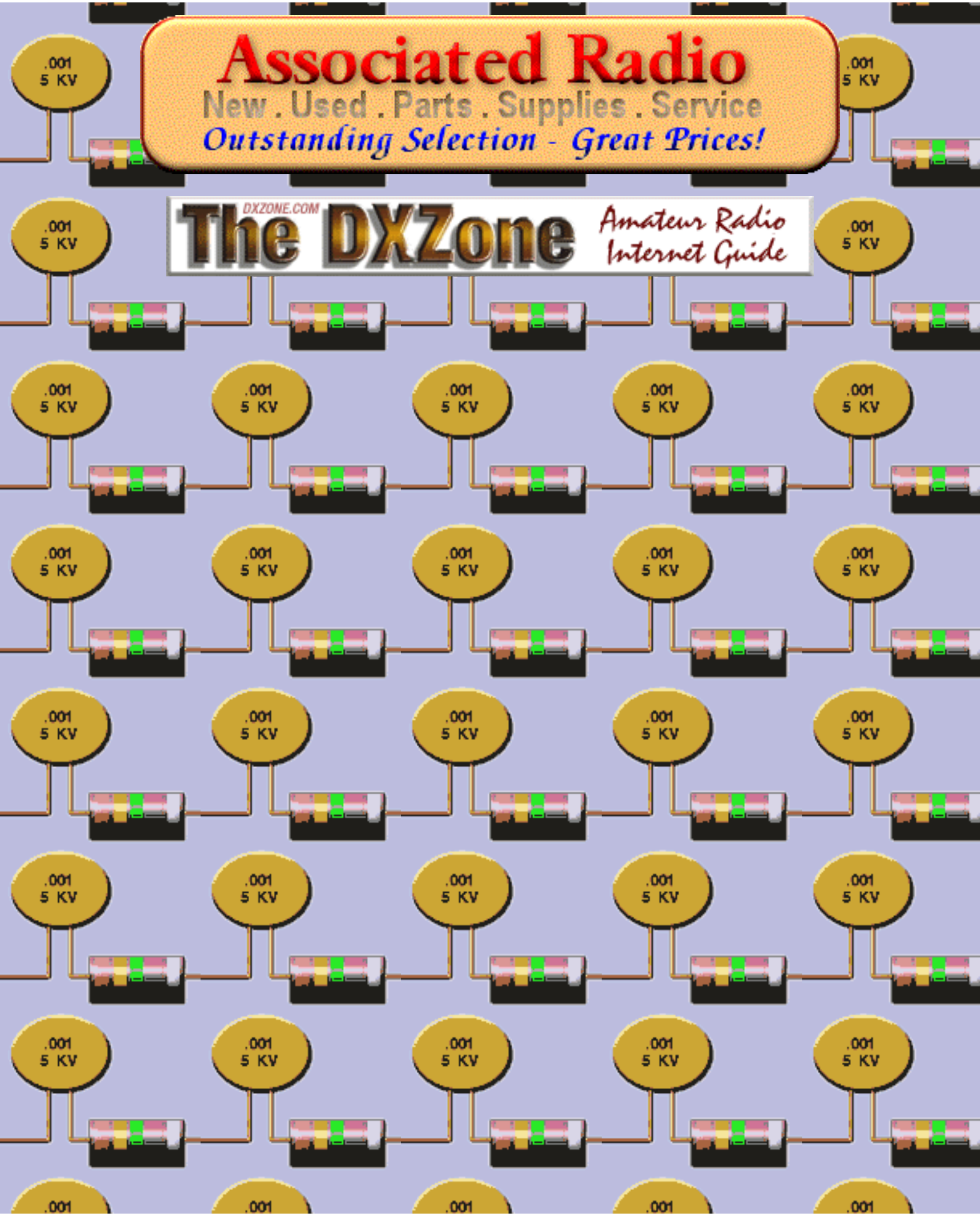
LEGAL LIMIT AMP PROJECT

established endeavor - QSLWorks - where we design and produce full color, custom QSL cards as a means of supporting this site, it's a "no digital zone", with absolutely no attention given the concept of political correctness. For that reason, the theory, practical experience and simple humor is entirely analog.

For those new to ham radio, a special welcome is extended to you! Here, you will find years of practical experience offered simply for the taking; the kind of experience gained by hams that fearlessly applied theory and the use of available (often scrapped) parts and supplies (with mixed success and occasionally shocking results) to the design and construction of a variety of station accessories. If you are just getting started and are looking around for equipment, we've added a new section to help you get off on the right foot. Your First Rig provides info on a number of older rigs on the market that will get you up and running with a great signal, and without breaking the bank. Once you know what you're looking for, you might check out Associated Radio. They're a fine bunch of hams with a great line of new and used equipment, as well as many hard to find parts and supplies. I've personally done business with them and was very pleased with the personable service they offer. In Parts is Parts, you'll find information on salvaging parts and supplies, as well as articles on a variety of ideas such as how to wind your own chokes, and creating custom meter faces. In Reference Material you'll find an assortment of resource material that will be of use if you're seriously interested in home brewing equipment and accessories. And, if you're looking for the highest quality, full color, custom QSL Cards, then check out our QSLWorks! We established the business as a way to help support this site and are currently the only QSL printer offering premium, full custom, full color QSL cards in the US. According to the feedback we've been receiving, we are also very likely one of the Best.

So, grab a fresh cup of coffee (or stick with the one you've got), sit back and enjoy. Check out the tutorials, projects, and ideas that are featured. And, if, in the process, you think of something you'd like to see added, or perhaps make a comment on the site, or even better yet want to meet any of us on the air, just drop us an e-mail at your leisure.

Best 73s - Bob WB0NNI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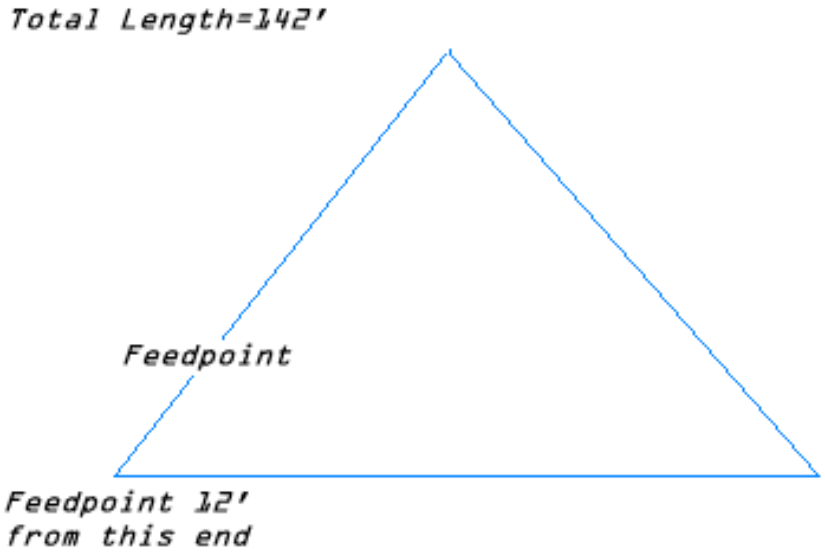
- K4PX
- AF4Z
- WB4EQS
- WT4I



- **Contruction:** I made and put this antenna up the Friday afternoon before the ARRL RTTY Roundup (1999)! It took me about 3 hours to make and install on the tower. Most of that time seem to be used untangling the wire..hi hi

- I made the antenna 142 feet long. Use tiewraps at each of the three points to form a small loop at each end so that A 3/16" rope could be used to support the antenna. I used a center insulator that I had for a dipole--it already had a SO-239 which made it easy to attach the coax feedline. The feed point was

- KC4HW

12' from the bottom of one leg.

- This antenna requires a matching element of feedline made out of RG-59 (75ohm) coax. The formula for the length is a 1/4 wavelength at the operating frequency times the velosity of the RG-59. The RG-59 I had, had a polyprolene dialectric therefore the length turned out to be approximately 22 Feet. It was Easy to figure, a 1/4 wavelength at 7.050 (234/7.050) = 33.19' times 66 percent equal 21.9.

- Since I had the center insulator with the SO-239, I put PL-259s on each end of the matching coax and then used a bullet splicer to put the regular 52ohm coax feedline to the antenna..

- Because my mounting point was only about 42' above ground, which was insulated from the tower by a 3' fiberglass pole, I move the legs of the antenna out from the bottom of the tower approximately 25' to give a clearance under the antenna of about 7'. The legs of the antenna were secured East and West, and North and South was broadside to the antenna.

- **Results:** Actually, the antenna played better than anything that I have had previously. The SWR was flat from 7.0 to 7.1 only rising to 1.7 at 7.3. This was perfect because I wanted it to play RTTY and CW... I did find that the antenna appeared to be less effective broadside, to the North. Stations in New England were difficult to get, however, the Europeans and Western USA were answering on the first call. Also, a few days after the contest, I talked to K4HXW/Mr Tucker, he indicated that he thought the antenna was playing much better than the inverted Vee that I had previously.

- **Conclusion:** I believe that this makes a great DX antenna. Also the antenna did work on 15M. I did not try it on any other bands, but it should work well on 10 thru 40 with a tuner.. Making it a nice choice for a second radio antenna during a contest... Hopefully, I will know more about it in future contests... cu in the pile!

  Added 31 Aug 2002 Update: I have put this antenna back up recently and found that the correct length was a little off on my original antenna. According to ON4UN's Low Band DXing book, the proper length for a delta loop should be 1.05 to 1.06 wavelength. The correct formula to calculate this length in feet is to divide 983.59 by the frequency in Mega Hertz (983.59/7.150=137.56'). I also have changed where the antenna is feed to horizontal by feeding the antenna in the middle of the bottom. This is so that I could have better effectiveness in domestic contest. After several days testing, the antenna seem to work well. Only time will tell... hi hi

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K4TX

Chester, Virginia

Welcome to K4TX's small bit of "Cyberspace!"

LOOK !!!

PY4VE's 2 Element 40 Meter Delta Loop Photographs!!!

CLICK HERE à PY4VE

Hi - my name is Chuck Stigberg, licensed since 1963. I've had a number of calls including:

WN4QIT - WA4QIT - KA8HLZ - KB8TA - KD4FP - NT4U - K4GE - K4TX

My primary interests are SSB, CW, Contesting, QRP, Construction & Experimenting with Wire Antennas - and yakking with my good friends on 3.868 Mhz. I look forward to meeting "you" on the bands.

- **EQUIPMENT**

Yaesu FT-1000 MP.

Ameritron AL-1500

NYE Viking MBV-A

Kenwood TM-255 E Multimode 2 Meter Rig

ANTENNAS

160 Meter Inverted L - Vertical portion 70 Feet - Radials (need work!)

75 Meter Horizontal Dipole @ 65 Feet

40 Meter 2 Element Parasitic Delta Loop - Switchable NE/SW - @ 60 Feet In Pine Trees

2 Meters - Diamond - Base is @ 80 Feet in Pine Tree

DESCRIPTION OF K4TX DELTA LOOP

The Delta Loop array used at my QTH is a simple and inexpensive way to achieve substantial gain and reasonable F/B ratio. I took the idea from Dave Pietraszewski, K1WA's article in the ARRL Antenna Book when he wrote about 5 Sloping Dipoles suspended from a single tower. The sloper parasitic array he described had an "ungrounded" coax switch mounted in the tower. He used 3/8 electrical wavelength feedlines to each element. He would transmit with a single sloper & the other ones acted as reflectors. Why? Because the 3/8 wavelength feedline going to the other slopers (each being open circuited due to the ungrounded coax

switch) added inductance making the element appear 5 % longer - thus acting like a reflector.

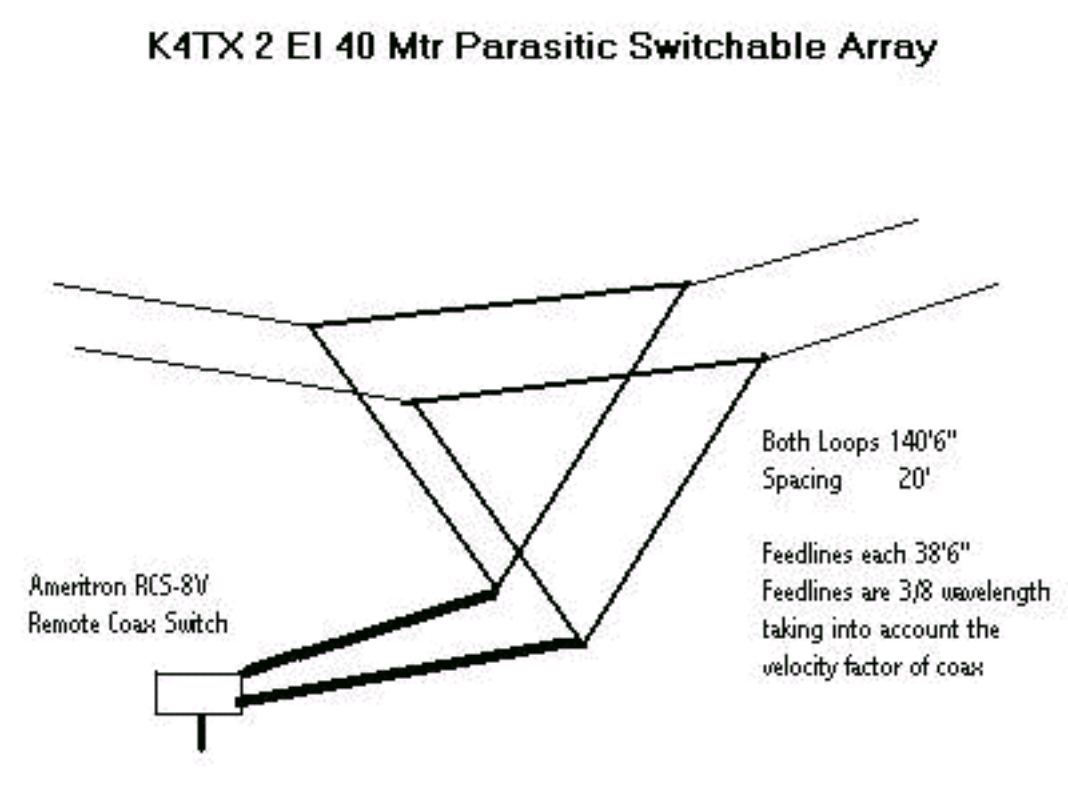
I decided to see if it would work with a pair of equilateral loops. I already had one up, and I was disappointed with the performance. I had tried all sorts of things (i.e. switching the feedpoint to change polarization, inverted, sideways, you name it. . .to no significant avail.)

I put up the second loop - same size - spaced about 20 feet apart. It was no easy feat using my slingshot to get the ropes in the pine trees and maintaining some semblance of symmetry. I cut the feedlines (RG-8x) to the proper length. The velocity factor of the RG-8x was .81 according to the manufacturer.

Lastly - The switch. . .Ameritron's RCS-8V comes from the manufacturer with the coax ports "ungrounded." Voila! I fed the loops direct - NO quarter wave matching transformer needed!

PERFORMANCE RESULTS

Absolutely phenomenal!! On air reports confirm that this antenna is equal or better to most 2 element yagis and competitive with most 3 element yagis - albeit at the same height. And - you have to remember that "average height" of my Delta Loops are only 30 feet. The top is at 60 feet and the feed point is only about 15 or 20 feet. Typical reports? VK/ZL usually give me +20 db over S-9. I can also work them long path. Europeans typically give me the same reports or better. The front to back ratio is only about 10-15 db - but what a difference it makes to cut the European broadcast QRM when I work stations to the West! They literally go from unreadable to solid copy. Best of all is the antenna is very inexpensive and INSTANEOUS SWITCHING.



FAVORITE LINKS

NOAA WEATHER

AC6V - The best for ham related links!

GI0AIJ - "Super Station!"

The Nerd Net

Virginia QRP Society

Sons of Confederate Veterans

PERSONAL INFO

I'm 52 years old, and still enjoy the hobby a great deal. It's difficult to

fit all of my interests into my day to day living. These interests Include: Golf, Relic Hunting, American Civil War History, Writing, and a host of others too numerous to list.

I have an MBA degree, am a Vietnam vet, and work as President for a Cable TV Construction Company. I've worked for the various large CATV companies in the past, and I was even fortunate enough to begin my own cable TV system that was ultimately sold to a large company after only 3 years. That was back in late 1989. My company now provides turnkey aerial and underground construction of coax and fiber optic cable with services predominately in the Mid-Atlantic region of the US. I'm married to a beautiful wife. I have a 23 year old daughter, a 13 year old son – AND Twins were born November 17th, 2000. Sean & Shelby. . .a boy and a girl!

E-Mail K4TX

Amateur Radio Station K7ZB Arizona

Home
K7ZB QSL Card
A brief K7ZB history...
K7ZB activities...
160m DX Contests - 2001/2004
Jamboree On The Air 2000
A radio attempt on Kathmandu...
An effective 10-20m DX antenna for deed restricted lots...
The 40 meter Stealth Vertical
A Stealthy 75 meter Antenna
HF Receiving Loop Antenna
Other activities and photos
Email K7ZB

The 40 meter Stealth Vertical

As published in www.antennex.com Dec. 2001

"You're 30dB over 9 here..." So goes the consistently fine signal reports received from around the USA and beyond - on 40 meters at the peak of Sun Spot Cycle 23. The most common antenna used in ham radio mounted over poor desert soil conductivity still performs beautifully!

This is the view of our second floor deck as seen from the closest street. The need for a 40 meter antenna that would perform well and not violate the spirit of the Home Owner's Covenants protecting the aesthetics of the neighborhood was the driving force behind the design of this vertical.

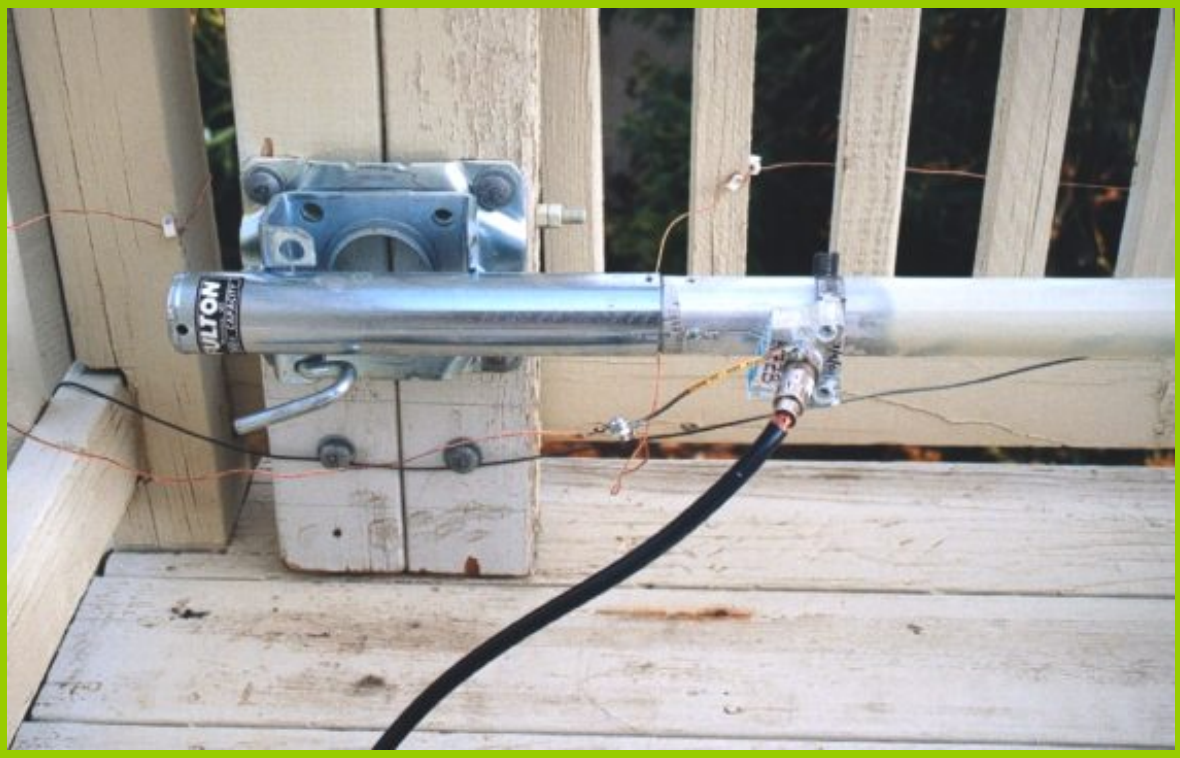
This antenna was designed to provide low angle radiation for good DX performance during the night time hours. DX on 40 meters is best when the local sun is down and this makes it convenient to use the cover of darkness to hide the size of a quarter-wave antenna. Especially one which is mounted 12 feet above ground which puts the top of the vertical at nearly 50 feet!

The basic concept is to mount a standard 1/4 wave vertical element on a swivel mount, secured to the deck railing. The mounting must be extremely secure when the antenna is in the upright position. It should also be easy for one person to put up or take down in less than 5 minutes. This design meets these criteria with excellent results.

The 40m vertical in its down position rests along the bottom of the far side deck railing. It is supported by plastic coated hangers of the type sold in hardware stores for hanging bicycles, etc, on garage walls. These hangers also make excellent supports for the antenna in the intermediate position for extending the telescoping top section and for supporting the vertical in the upright position.

Shown below is a view of the vertical with the telescoping top 5' section removed and stored in the down position. The swivel assembly has a spring-loaded locking pin which secures the base in either the upright or down positions. Putting up the antenna is simple: the top section is secured with a wing nut then the

vertical is walked up hand-over-hand into the upright position and locked in place with the pin. The hanger also stabilizes the antenna so it does not sway in light wind.



The swivel assembly I modified is a Fulton TJ503 swivel jack stand with snap ring, purchased at Walmart's. Fulton's website is available here.

The swivel assembly has been modified to support the vertical base element. Its load rating is well in excess of the load imposed by the vertical. Since the 35' - 7 1/2" vertical is only intended for use during good weather conditions it is only guyed with two light guy lines to ensure that in case something did happen to the antenna while up, it will not fall across neighbor's property. In our neighborhood there are no above-ground wires for power, cable TV or telephone, so there is no possibility of a crossed-wire mishap.

The coax attachment is made through an SO-239 connector mounted on an acrylic plastic block drilled and U-clamped to the base tubing. Also visible are the two radials connected to the shield of the coax connector. The two radials are 33' long, and slope from 12 to 7 feet above ground at their end. They slope because that's the available tie point height in the yard. The radials are oriented 145 degrees apart - not quite the 180 degrees desired but close enough. The EZNEC antenna azimuth plots do show the minor skewing of the pattern due to the asymmetric radial placement, but this has little effect on its performance.

The net active dimensions (not including the length inserted into lower elements) for each element of the vertical are as follows:

Base element: 11' 5" (1 3/4" Dia. Al tubing)

2nd element: 10' 2" (1 1/4" Dia. Al tubing)

3rd element: 5' 7" (3/4" Dia. Al tubing)

4th element: 4' 11 1/2" (1/2" Dia. Al tubing)

5th element: 3' 6" (3/8" solid Al rod)

The tubing diameters were based on what I had available. Good mechanical design technique should be used in attaching each element securely into the lower one.

This antenna does bow substantially when being put up and down. This droop could be minimized by going to a higher strength alloy. The best strength-to-weight ratio for vertical tubing is probably titanium-aluminum alloy, although it costs substantially more than the material I used.

Raising and lowering 35 feet of aluminum tubing up over one's head in low light conditions leads to safety considerations. I wear a hard hat and safety glasses when raising or lowering the vertical.

A Word document is available which contains a few more construction details and EZNEC plots for this antenna. Send me an email and I will provide a copy.

Shown below is a night-time photograph of the vertical in the upright position. The stabilizing hanger is seen approximately 2' above the swivel assembly. Notice a 6 turn coax loop in the line which serves to help keep RF out of the shack - which is about 20' from the vertical. The RF exposure on 40 meters with 500 watts output is within the FCC's Maximum Permissible Exposure limits.

An advantage of verticals mounted above ground like this one is the safety aspect of proximity to RF-hot radial wires or vertical elements. Our yard is walled and the RV gate kept locked, and the access out to the deck is past my operating desk - so there is little danger of anyone's unexpected exposure to hot wires.

Here is Brian (KD7Z) looking up at the top of the vertical with Dave (AB7E) out on the deck.

Brian helped with the design concept when he directed my attention to a Hints & Kinks article in QST (May 2000, page 56) that featured the fold-down mobile-antenna mount design of KB5YA. It was this use of the swivel jack that led us to conceive its use for a deck-mounted 40m vertical design.

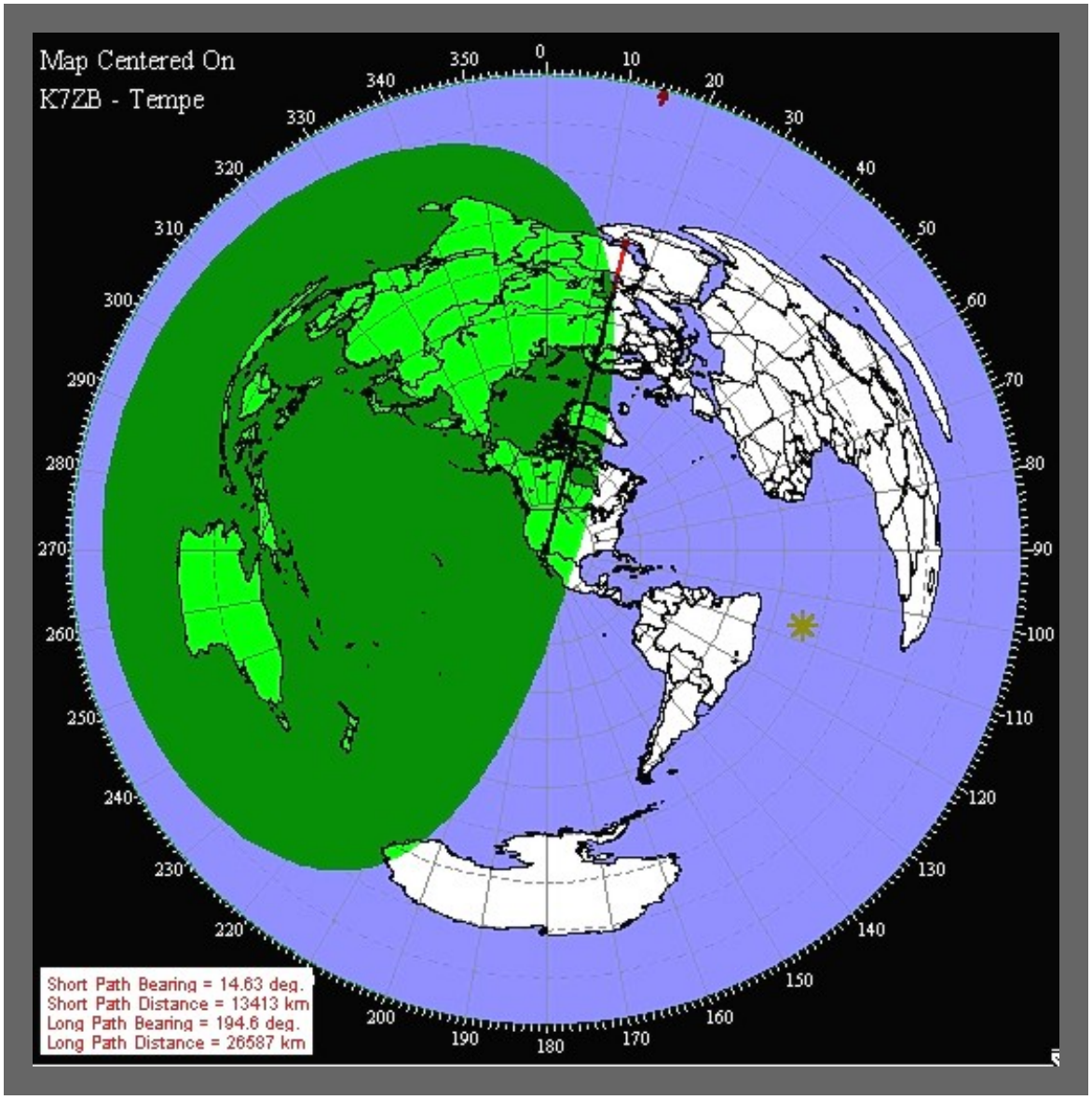
Dave provided EZNEC computer simulations for the vertical dimensions used here, which when implemented were right on target.

The photo below is of the vertical in daylight - up just long enough to take the picture to show how it looks at night! You can faintly make out the guy lines and one of the radials running out to the spot where I was standing to take this picture. Above the doorway on the deck is the coax feedline to the center insulator at the eave under the roof peak for the 75m antenna.

An excellent reference for understanding vertical antennas is provided by L.B. Cebik, W4RNL. Dr. Cebik is an authority on antennas and his website contains a vast amount of excellent information. The webpage specifcally covering verticals is shown below.

"Verticals Without Vertigo" by L.B. Cebik, W4RNL

Here's a QSO I made with A61AJ in the United Arab Emirates on 40m CW using this antenna and 100 watts. The Great Circle Bearing map below (AZ Map by Paul Burton AA6Z) shows the conditions. Gray-line propagation makes contacts like this possible and a vertical antenna can help you take advantage of it.



Signal path analysis:

This QSO illustrates a typical ***Long Path*** communications link. It is possible that the path was established with multiple hops of our signals in both directions. RF leaving my antenna would have launched at the low take-off angle associated with vertical antennas then skipped off the F2 layer of the ionosphere and returned to earth somewhere near the maximum single-hop skip distance of 4,000 km. This would place the first approximate reflection point out over saltwater on a bearing of about 195 degrees South of my QTH and in the South Pacific. The excellent conductivity properties of saltwater would allow efficient rereflection of the signal back up at a low angle and the process repeated for about 6 hops when the RF finally arrived from offshore into the United Arab Emirates. Other propagation modes like chordal hopping could have been involved as well. The enhanced propagation due to the sun's terminator made the path work with a mere 100 watts

at my end. The picture below was taken of this sunrise just minutes after the QSO.

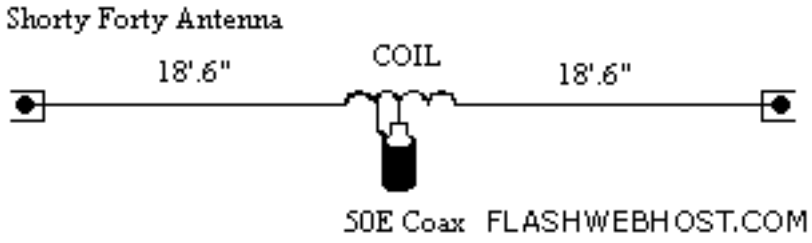
The sunrise at the time of the A61AJ QSO...

Both 40m and 20-10m verticals are up. Note the stars still visible so early in the morning!

Shorty Forty Antenna

City dwellers and other flat dwellers know the practical difficulties they encounter when they try to put up an antenna for 40 meters. Finding space enough to accommodate 67 feet of wire is a real problem for them and most of them have to be satisfied by the inverted V. Whatever the book say, in practice the inverted V does not come up to the standard of a dipole antenna.

The Shorty Forty Antenna can be put up in the space required for a 20 meter dipole. If is a compact 40 meter dipole for limited space application by Jact Sobel W5VM. It is a center loaded antenna with a loading coil at the center. The two arms are 18 feet 6 inches long connected to the two ends of an inductor at the middle. The inductor consists of 30 turns of 12 SWG enamelled copper wire wound on 2.5 inch diameter PVC tube 5 inches long. There is six turns per inch so 30 turns will require 5 inches long. The shield of 50 ohms coaxial cable is connected to the center of the coil. The center conductor is connected to 2 or 3 turns away from the center, to a point which gives the lowest SWR. Compare this to a 67 feet dipole the saving in space is substantial.



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The Beam/Yagi Antenna

ARRL Technical Information Service page · TIS Menu page

Additional information on this subject and related topics can be found in back issues of *QST* and the following:

The ARRL Antenna Book
Physical Design of Yagi Antennas

The ARRL has an extensive catalog of books and materials related to Amateur Radio.

Articles

Note: Some of the following articles are in Adobe Portable Document Format (PDF) files. To view and print these files, you'll need a copy of Adobe's Acrobat Reader program. (Version 3.0 or later required). More information here.

- Simple Offset Feeding of Wire-Element Beams (209,735 bytes, PDF file) MEMBERS ONLY
  QST October 1999, pp. 45-46
  This approach to matching a feed line to an antenna uses the antenna itself as an impedance transformer.
- Why A Beam Antenna? (1,221,309 bytes, PDF file)
  QST January 1972, pp. 36-39
  Some basic antenna information for the newcomer about Yagi antennas including a tutorial on antenna gain and construction of a 15-meter beam antenna.
- Simple Gain Antenna for the Beginner (778,544 bytes, PDF file)
  QST August 1981, pp. 32-35
  A tutorial on the Yagi antenna with construction of a two element beam for 10-, 15-, or 20-meters.
- The Building-Supply Yagi(920,911 bytes, PDF file) MEMBERS ONLY

ARRL Products: Interference/DF
(More)

AC Power Interference Handbook -- New insights into the causes, effects, locating and correction of power-line and electrical interference. 2nd Edition.

Transmitter Hunting -- Radio Direction Finding Simplified

The RSGB Guide to EMC -- Tackle RF interference problems and understand the underlying causes.

The ARRL RFI Book -- Practical Cures for Radio Frequency Interference.

QST March 1991, pp. 22-24
Here's a cheap, easy-to-assemble, two-element Yagi you can build for 10, 12, or 15 meters.

- Two on 10 (237,149 bytes, PDF file)
  QST April 1999, pp. 67-69
  A two element 10-Meter beam designed for portable or permanent installation. (Additional information)
- A Two-Element Duoband Beam (880,439 bytes, PDF file) MEMBERS ONLY
  QST April 1993, pp. 36-37
  Explore the 12- and 17-meter bands with this small, lightweight Yagi.
- A 15-Meter Beam On A Budget (602,286 bytes, PDF file) MEMBERS ONLY
  QST February 1971, pp. 41-43
  A two element beam made from electrician's thin wall tubing.
- Basic Beams for 12 and 17 Meters (1,494,442 bytes, PDF file) MEMBERS ONLY
  QST August 2000, pp. 57-62
  Some well-designed and easy-to-build antennas for the 12- and 17-meter bands.
- A Three Element Lightweight Monobander for 14 MHz (178,829 bytes, PDF file) MEMBERS ONLY
  QST July 2001, pp. 28-31
  A portable easy to build light weight antenna
- A Portable 2-Element Triband Yagi (257,252 bytes, PDF file)
  QST November 2001, pp. 35-37
  This novel wire antenna is great for permanent or portable, QRO or QRP, and old-timer or beginner operation.
- Practical High Performance HF Log Periodic Antennas (1,238,771 bytes, PDF file) MEMBERS ONLY
  QST September 2002, pp. 31-37
  The Electrical and mechanical design process for two Log Periodics that cover the HF bands from 10-30 MHz.

Note:

Contact information for suppliers mentioned in the above articles should first be confirmed using **TIS Address Database Search**.

Bibliography (Members Only)

ARRL Periodicals Index Search - This database contains the *QST* index from 1915 to the present and the *QEX* index from 1981 to the present. For *QST* issues from 1970 to the present, and some selected articles back to 1922 (when construction articles featuring tubes began in earnest), identifying keywords have been added to the technical articles. By entering keywords (ANTENNA) or combinations of keywords (CONSTRUCTION ANTENNA VERTICAL HF) into the **Title words:** field, you may create dynamic bibliographies.
Technical article KEYWORD list. Hints for more successful searching

Suggested keywords for more articles like the ones on this page are: CONSTRUCTION HF BEAM ANTENNA

Web Links:

- K5TR
  A collection of antenna modeling files

- Mini-Beam Geometry Basics
  There is a natural progression of the geometry of the "mini-beams" deriving from the basic yagi, which in itself owes a lot to the common dipole.
- Stacking Yagi Antennas
- KB0YKI's Radio Zone
  On-line antenna designers, rooftop tower plans, gamma matches, and more
- DXCC Country List / Beam Headings
  Display a current DXCC list with beam headings centered on your location.

Article Reprints:

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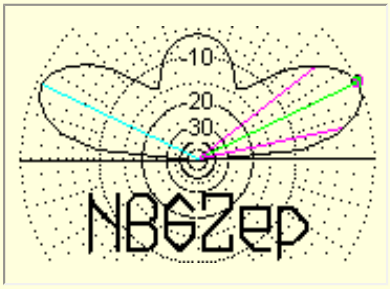
NB6Zep Antenna

Basic Design: A modified 20 meter Extended Double Zep ***Operating Bands:*** 40 thru 10 meters (with tuner)
Dimensions: 66 feet horizontal, 5 feet vertical ***Transmission Line:*** Open Ladder or TV Twin Lead
Materials: Conventional low-cost wire antenna construction (see below for details)



- The Super Zep is an elegant multi-band wire beam found **here**.
- Build a wire dual band DX Dipole or Yagi antenna system **here**.
- Instructions to add 80 meters to the NB6Zep are found **here**.
- Instruction for the smaller NB6Zep Jr are found **here**.
- A Space Saving 80 meter antenna is found **here**. New

The NB6Zep antenna is a simple inexpensive wire antenna, easy to build, that will yield excellent results on all bands. Only a dipole would be easier to construct and maintain. Unlike a dipole, the NB6Zep will operate on all ham bands 40-10 meters (including the WARC bands) with no strain on the antenna tuner, transmission line or operator. When you follow the basic rules for this antenna and use good common sense construction techniques to build this antenna, you will receive reliable results as shown in the documentation I have provided here.

> *"I wanted a portable antenna I could throw up between trees while camping so I made the NB6Zep on your web page. To test it at home I (naturally) set it up between a tree and a old section of TV mast tied off to the field fence. It's all of 15' off the ground but is noticeably better when I A-B compare it to a G5RV at 35'. The antenna is still up proving there is nothing so permanent as temporary when it works!*
> *Thanks for putting the design on the web. Easy to build and the darn thing works..." Jim Edmondson WA7KQG*

I live in an area with severe antenna restrictions. Not even TV antennas are found on the houses in my neighborhood! For the last eleven years I have been operating successfully on the HF ham bands by hiding my "evil activity" in the pine and fir trees that line the back yard of my home in Beaverton, Oregon. I have experimented with many designs and techniques for low-profile wire antennas over the years and have learned what works and what does not work. I have 5 wire antennas integrated into the back yard, but the single most valuable antenna I have is the NB6Zep at 40 feet above the yard! An NEC wire modeling program is needed to keep all of my wires operating properly in the space of a typical back yard. I have spent many hours using EZNEC to model the NB6Zep antenna. The total length is optimized to fit in the space of a 40 meter half wave dipole and the feed line length is optimized for good operation when attached directly to a tuner with balanced output. A coax termination to the tuner can be used with a 4:1 balun device at the end of the balanced line. I am happy to share with you the details for building your own NB6Zep.

Basic Construction: The wire must be installed as horizontal as possible. Do not invert or "dog leg" the wire as this will impact expected performance. The wire and feed line are light enough that it can be mounted with just the two ends supported. A strong 18 AWG wire is used to make both legs of the center fed antenna. Cut each leg exactly to the same length of 38 feet. (33 feet horizontal and 5 feet to hang down vertical.) Use a heavy duty ceramic or plastic insulator for the center to connect the wire legs to the feed line. ***CLICK for DETAIL*** The modeled feed line

length was 49 feet long for open ladder line and 41 feet long for TV twin lead.
Use a medium duty ceramic or plastic insulator at both ends of the horizontal wire to serve as a tie off point for the antenna support and to support a 90 degree bend to drop the last five feet down vertically. ***CLICK for DETAIL*** Attach a medium size lead fishing weight at the end of the wire to keep gravity working for you. Use a section of nylon twine (string) to hold back the vertical wire from wrapping itself around the support mast or other objects when the wind blows. ***CLICK for DETAIL***
Mount the NB6Zep as high as you can manage to support it. Drop the feed line down or away from the antenna at 90 degrees for as far as possible. Keep the feed line away from metal objects if possible. The line may be attached directly to the balanced line output posts of any tuner. ***CLICK for DETAIL*** If that is not possible, construct (or purchase) a 4:1 voltage balun and connect the step-up balanced side of the transformer to the twin lead and connect a short run of coax to the unbalanced side. You connect the other end of the coax to the unbalanced output connector on the tuner. This technique is useful for bringing twin lead from the outside into the shack through a window or other entry. Make a loop at the balun transformer to keep rain out of the device. ***CLICK for DETAIL***

Basic Performance: While waiting for the rain to stop so that I could install the antenna at 40 feet, I operated for several days with it stretched between two trees at 6 feet off the ground. Good psk31 contacts were made from Oregon to the East coast with 50 watts on the 40 and 30 meter bands. (High angle radiation is very under rated on the low bands). The antenna was brought to 40 feet, supported by two light gauge steel Radio Shack masts strapped to the upper trunks of two pine trees about 70 feet apart. The antenna is broadside to the SE and NW. All of the design data from EZNEC is based on the NB6Zep at 40 feet off the ground and with no large objects or other tuned antennas in the near field. The gain of the NB6Zep is based on the maximum field from a 1/2 wave dipole at a 1/2 wave length above ground. Your performance will differ if, for example, you have dipoles, yagis or any horizontally tuned antennas in the same yard as the NB6Zep. Your results may differ if your antenna height is less than 35 or greater than 45 feet. You should consider removing other horizontal antennas as they will be mostly obsolete with the NB6Zep installed. If possible, align your NB6Zep in the direction that places lobes where they will reach the largest population of ham operators. Nulls in the antenna above 30 meters can be useful to reduce QRM and interferences sources that are picked up by the horizontal polarity of the NB6Zep.
The NB6Zep should perform well at lower heights, even down to 20 feet. The take off angle will be effected by becoming lower at heights above 40 feet and higher for lower construction. It is possible to have good operation with the legs inverted down from the center feed point, however, the diagrams provided here are not valid unless the antenna is horizontal with the ground and in a straight line.
The NB6Zep is not a "tuned" antenna and will not interact significantly with other antennas in the yard. You could construct several NB6Zeps to have selectable directivity for the bands above 30 meters.
40 and 30 meters: The NB6Zep performs like a half wave dipole, with broad, medium to high angle lobes running off the sides of the antenna. Very good for regional and for long distance skip. The antenna displays very little directivity on the low bands and is performing very well in all directions.
20 meters: On this important DX band, the NB6Zep has it's highest gain, over 3 dB. There is a sharp medium angle lobe running 90 degrees from both sides of the antenna. ***CLICK for DETAIL*** For the first few contacts on 20 meters with the NB6Zep, I reached Texas, Utah, Alaska, Korea, Japan and Argentina, all with strong reports. A good QRP contact to Japan was made on PSKHELL mode. ***CLICK for DETAIL*** The antenna produces very sharp lobes from each side, good for long distance and regional skip.
17 and 15 meters: On these bands the NB6Zep produces two broad lobes spaced 45 degrees from both sides of the antenna; each lobe has gain. This looks like a butterfly pattern when viewed from above the antenna. ***CLICK for DETAIL*** The antenna performs very well for long distance with it's low angle radiation and also has medium angle radiation for shorter skip operation.
12 and 10 meters: On these higher bands the NB6Zep produces two narrow lobes with gain, separated by 60-80 degrees from each other on both sides of the antenna in an "X" configuration. ***CLICK for DETAIL*** Two layers radiate from the antenna giving very good operation for long skip and shorter skip conditions. As on 20 meters, the NB6Zep is very directional on 10 meters.
A diagram showing the current distribution on the NB6Zep antenna system for the major bands is ***FOUND HERE***

Construction Notes: Ceramic "egg insulators" are commonly used in wire antenna construction, but I have had

good results over the years making my own wire antenna insulators from plastic coat hangers. (Use the flexible round type purchased in stores, not the free hangers from the dry cleaners.) They are very strong, UV resistant and low-cost. Cut 3 inch sections and drill holes a 1/4 inch from the ends. Use them at other lengths for wire spacers when making a "fan dipole" or open ladder feed line.

Always solder every wire connection and then insulate connections from water seepage. Also insulate around all exposed coax braid. Use "hot melt" glue or a silicon rubber sealant like RTV. ***CLICK HERE*** to see some of the basic materials and tools I use to construct wire antennas.

You can make your own light weight balun using Iron Powder Toroid cores. The smaller T130 HF material (red) cores work fine for several hundred watts. Follow the diagrams in the ARRL handbooks. Use a hand drill to twist 20 AWG enameled wire into winding sections used in bi-filar and tri-filar transformers. Make about 12 turns onto the core for good performance. ***CLICK for DETAIL*** Wrap layers of electrical tape around the balun to protect the core from damage. (Cores are like glass and once they are cracked the core permeability is altered.)

Use medium weight nylon twine for all of your wire antenna construction and support needs. It is extremely strong and lasts forever.

The NB6Zep Story: Once upon a time, in a land far to the Northwest, there was a full sized 20 Meter Extended Double Zep antenna supported only by two majestic pine trees. The trees both stood at the very outer regions and at opposite corners in the land known as NB6Z. This span of distance allowed the EDZep to spread, although the weight of the feed line at the center could not be supported, causing a rather unsightly and disconcerting droop to be seen. Despite the odd appearance of this stately antenna, it continued to serve its' master well for many years; with many fine contacts on all bands from 40 through 10 meters. It occurred on one winter's day, that a fierce storm blew down one of the two pine trees that had supported the EDZep. This caused much anguish to the master of the land, who immediately gathered his forces to find an acceptable solution to this dilemma. Several seasons of trial and error passed, but no resolve could be made. Finally, after much calculating and re-calculating, a new design configuration was found and a new support tree was christened. Much to the delight of the master of the realm, it was found that this new design had no droop to catch ones eye. Further joy was had when the on-air performance of the new antenna was found to be just as good as the original! Overcome with joy, the master named this new design *NB6Zep*, after the land in which it resides. He further decreed that the new design be shared throughout all the land of Hamdom and with all who seek to enter the Kingdom of Ham.

"Push-button Menu of Stuff"

20m Delta Loop.

This antenna is fed for vertical polarisation, to give a low angle of radiation for DX and also a nearly omni-directional radiation pattern (broadside is the best direction by about 3 dB).

I have built two versions of this antenna; one using coax as a monoband antenna for 20m as simulated below, and the second as presently used with 450 Ohm slotted ladder-line feeder for multiband use (20 and 15m).

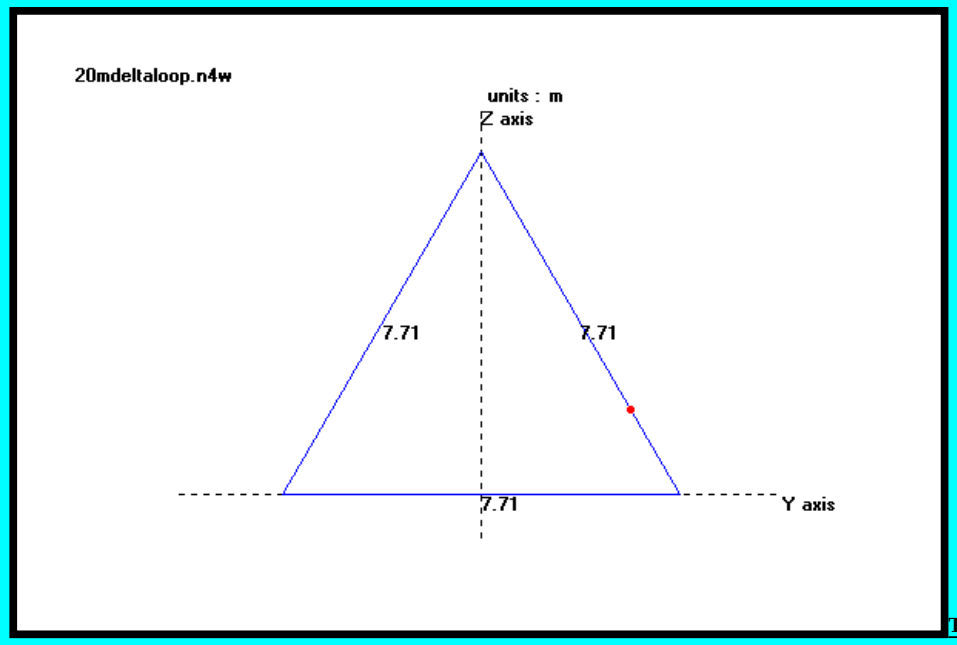
In the monoband application, the impedance at resonance on 14.2 MHz of about 100 Ohms is transformed by a quarter wavelength of 75 Ohm cable (3.5m long) to give a good match (better than 2 : 1) across the whole band. The 75 Ohm coax braid is connected to the lower side of the loop element. I don't use a balun as the antenna is not balanced, due to the offset side feedpoint, but do implement a coaxial choke at the 50 Ohm cable connection to the matching stub of 10 turns, 20cm diameter. The antenna is 7.71m per side, with the feed point 3/4 of the way down one side (5.78 m from the top) note; 3/4 * 1/3 wavelength = 1/4 wave, this is important to obtain vertical polarisation.

This antenna is also effective on 15m provided open wire feeder is used up to the antenna instead of coax. I have found 10m DX performance to be rubbish! Simulation confirms this.The pattern is rather directional on both the higher bands.

Photo of the antenna

TBD.

Drawing of antenna

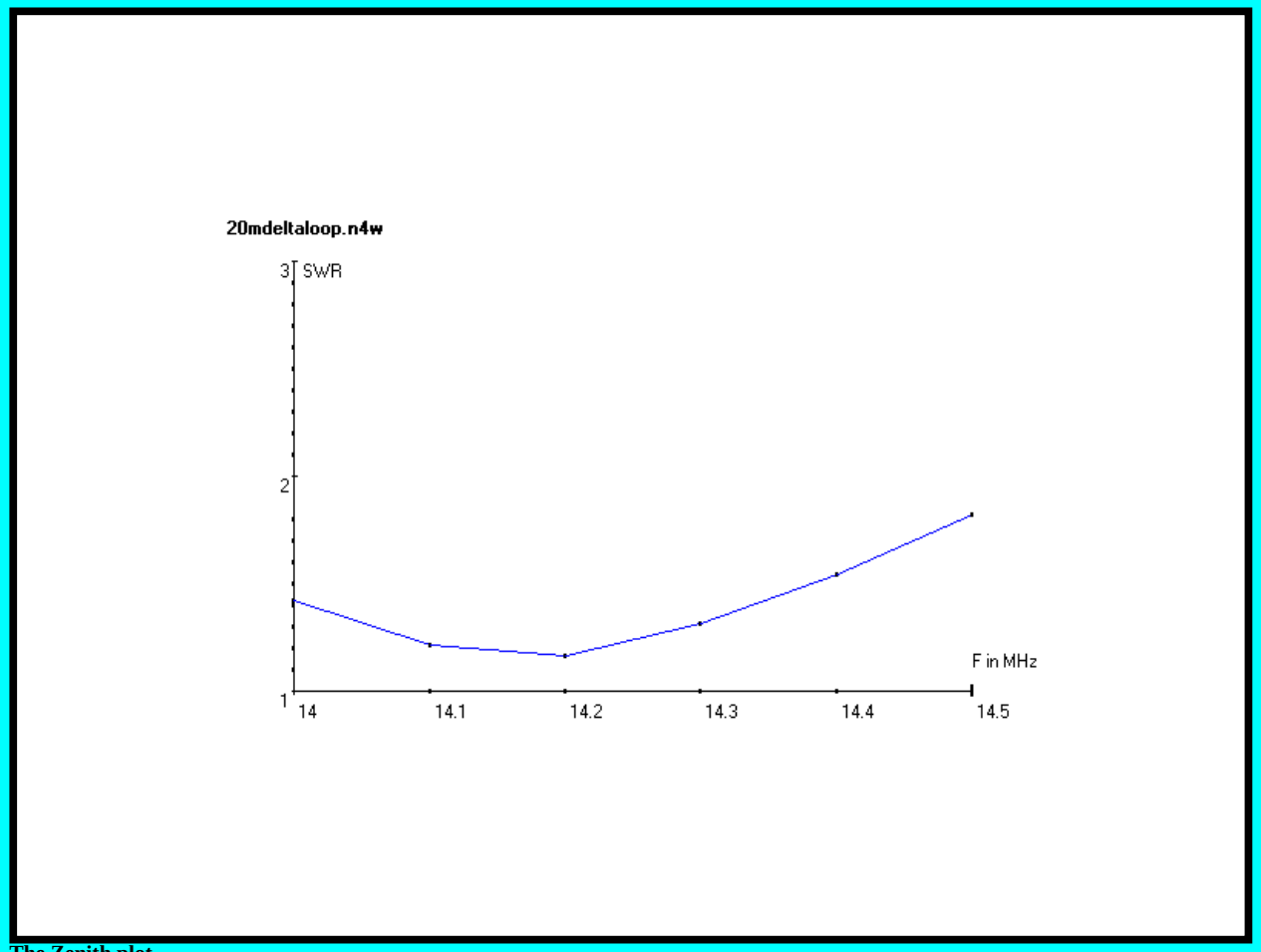


The NEC4WIN95 design file;

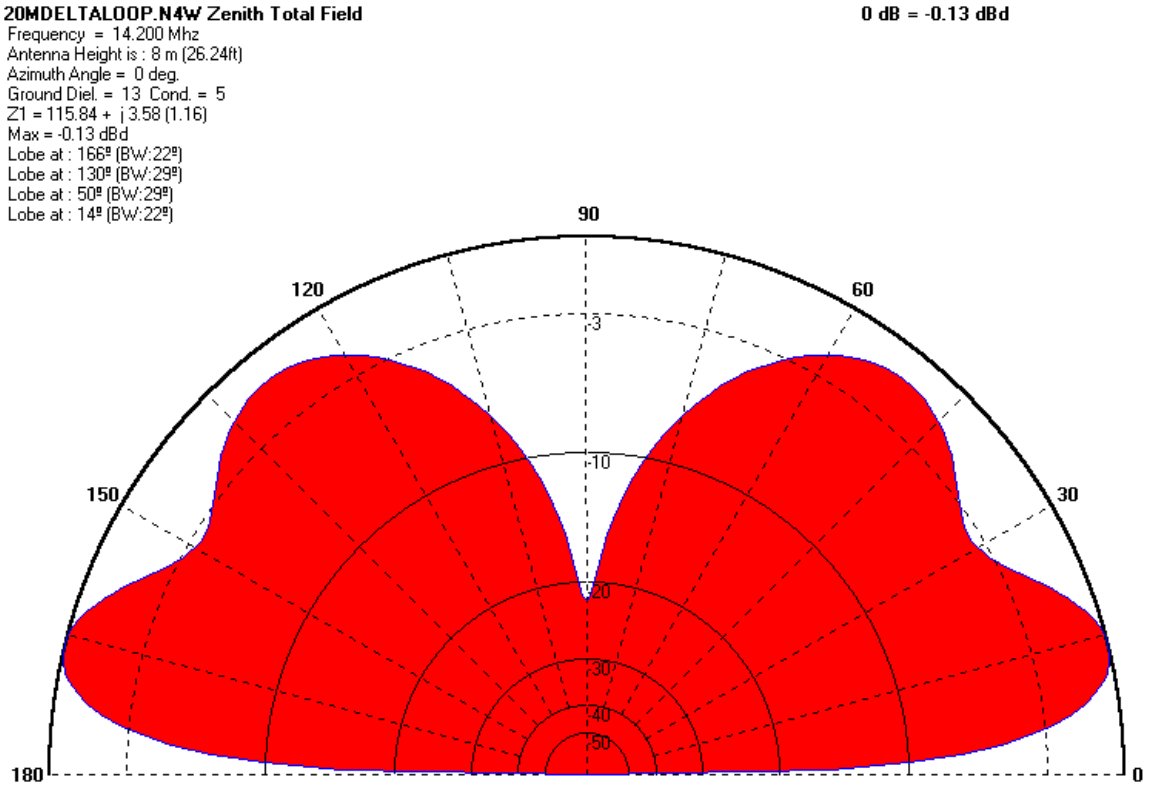
CM Equilateral Delta loop for 20 meters, apex up.
CM Feed point 0.25 lambda from top corner for vertical polarisation
CM 2:1 75 ohm coax quarter wave-length match assumed.
CM G4EZT Tree supported antenna.
CE
GND Reference
UNITS Meters
Height 8.000
Over Ground 13 5 (Diel. - Cond. µSiemens)
Boundary Circular
F 14.200

```
GW 0 8 0.000 -3.854 0.000 0.000 0.000 6.675 0.002
GW 1 20 0.000 0.000 6.675 0.000 3.854 0.000 0.002
GW 2 8 0.000 3.854 0.000 0.000 -3.854 0.000 0.002
S 1 23 100 0
Coax 100
```

The SWR plot



The Zenith plot

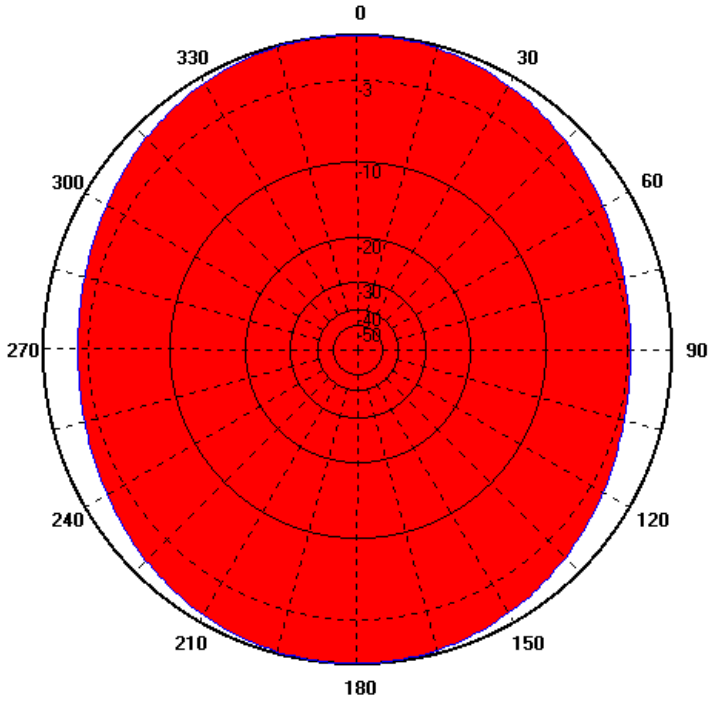


The Azimuth plot.

20MDELTALOOP.N4W Azimuth Total Field

0 dB = -0.12 dBd

Frequency = 14.200 Mhz
Antenna Height is : 8 m (26.24ft)
Ground Diel. = 13 Cond. = 5
Z1 = 115.84 + j 3.58 (1.16)
Zenith Angle = 14 deg.
Max = -0.12 dBd
F/B = 0.02 dB



A ONE ELEMENT V BEAM FOR 15 METERS

"I REFER TO THIS AS MY ONE ELEMENT BEAM, IT'S A HORIZONTAL V!
IT OUTPERFORMED 2 ELEMENT BEAMS WE HAD ON 10 METERS AND 20 METERS IN LAST YEAR'S FIELD DAY!
IT HAS GAIN IN A SINGLE ELEMENT DESIGN!

THE db GAIN, IF FIGURED BY THE BOOK, IS 3db OVER A DIPOLE!".....DARRELL REF 1
In log-periodic dipoles, it was found that forward tilt increased gain by a 3 to 4 db over a regular log dipole.
ref 2Horizontal V Antenna for 15 Meters
"The HORIZONTAL V ANTENNA FOR 15 METERS"
This often leds to the asumption of what is a Beam Antenna, in it smallest of states, the Horizontal Vee, must be one of the least researched or understood antennas, after stumbling on misleading data one would think the antenna to be a dud, that efforts to make such a antenna would be a waste of time. This is far from the case.....DARRELL.

This page is GRAPHIC intensive, please allow time to load!

A ONE ELEMENT BEAM FOR 15 METERS

CONSTRUCTION DETAILS

In log-periodic dipoles it was found that forward tilt increased gain by a 3 to 4 db over a regular log dipole. Ref2
A little history , 3 years ago, I was using a rotatable dipole for 15 Meters and had good results but wanted something better. This dipole was built with a DAK dipole mount and used 36 inch extenders with 6 foot 10 7/8 inch fiberglass CB whip antennas. This worked well, but I wanted something a little better. I spent more time reading about antennas and got interested in the Horizontal V antenna and found that the basic V starts at 1/4 wave with a 90 degree angle. Ref 1

Time to build and test my acquired knowledge.
The center of the antenna started life as 2 mirror mount CB antenna mounting brackets, the first piece was flat on both sides of 90 degree angle and a second 5/8 inch hole was added an equal distance on the other side.
The second mirror mount is heli-arced to the first piece, with the top piece horizontal and the bottom piece vertical with the pole mount going down. (**See picture above**) Welding was done at a local welding shop for a $5 bill.

Now I use the insulated CB 3/8 by 24 antenna mounting kits. You can pick these up at Radio Shack but get 2 and get the ones with bolts and stay away from the flimsy screws. The kits come with 2 bolts, 2 long nuts, and the plastic insulators. Use one kit per 5/8 hole on top, going from back to front (bolt, insulator, thru antenna bracket, insulator, long nut). Also while at Radio Shack pickup a couple of heavy duty terminal ends for the end of your coax. The coax is done in a pigtail fashion and connected on the bolt side of the antenna insulator hole 3/8 dia with at least a #10 wire connector for your terminal ends.
My elements consist of two 6 foot 10 7/8 inch fiberglass CB whip antennas with 38 inch extenders.(See picture above) and note that the original dipole used 36 inch but do to the induction between the elements, the resonant frequency raised to 21.350 MHZ. The 38 inch extenders lower the frequency to 21.250 with a l to 1 SWR). I operated field day on 21.205 and only got chased off frequency twice and came back and retook the frequency, minutes later. The extenders were made

from 1/2 inch EMT tubing. On one end a 3/8 by 24 long nut is inserted leaving 1/16 inch exposed for soldering. Soldering was done with a 5% silver solder. On the other end a 3/8 by 24 bolt, 1 1/2 inch long, with a 3/8 by 24 nut attached to leave 1/2 inch of threads exposed was inserted in the other end leaving 1/16 of the nut exposed and the 1/2 inch of threads. The nut is soldered on the exposed 1/16 inch of nut, take care not to weld on the exposed 1/2 of treads. Two of these must be made.

The final assembly starts at the center antenna mount. Connect coax pigtail on bolt side of antenna center insulators, connect treaded side extender to center, connect whip to extender and repeat for the other side. (**See picture above**)
The entire 15 Meter antenna weights around 6 pounds and was used on a 20 foot tower with an 18 inch truck wheel for a base, no guy wires were used and it withstood a 30 mile per hour wind when a thunderstorm hit just after setup for field day 2001. It has performed better then I could ever have imagined. I learned what it was like to be in a pileup and have fun, we even worked **DX** and had 49 of 50 states in the 24 hour period! We had directivity, and rotating from NE to NW we lost the **DX,** but started working the Western States. I figure the gain over a vertical at around 6 db, and a walk around the Horizontal V with a field strength meter (**see diagram above**) was enough to tell me that this antenna will perform.

A fellow ham came by, **AF4HZ** Gordon Blauser and I let Gordon walk around with the field strength meter and he walked away amazed with how good the antenna performed with 10 watts running to it. He got one of the proto type centers and was vowing to try using 2 hamsticks on his V. I also have to get Spencer Whitmire, **W4ERC**'s report on how another proto type center is being tested on 6 Meter with just a couple of stainless steal whips cut down to around 52 inches that will work on the Magic Band.

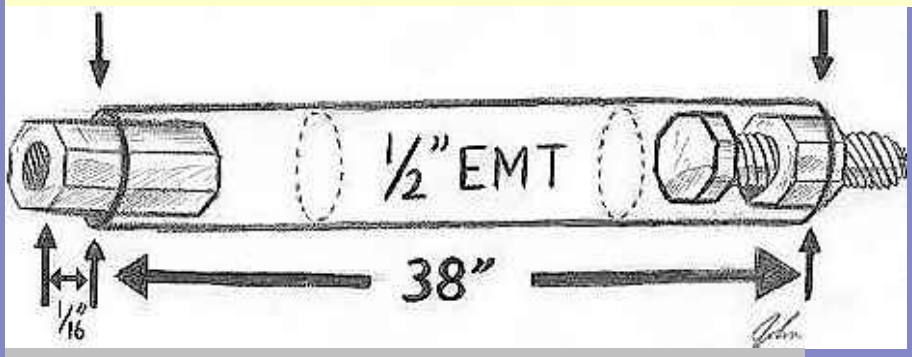
The center antenna mount could be made with a 6 or 8 inch longpiece of aluminum angle 2 inch by 2 inch and 1/4 inch thick. This should give plenty of space for the mast clamp. Nay sayers may say that the aluminum mast affected the radiation pattern but it worked for the good of the antenna.

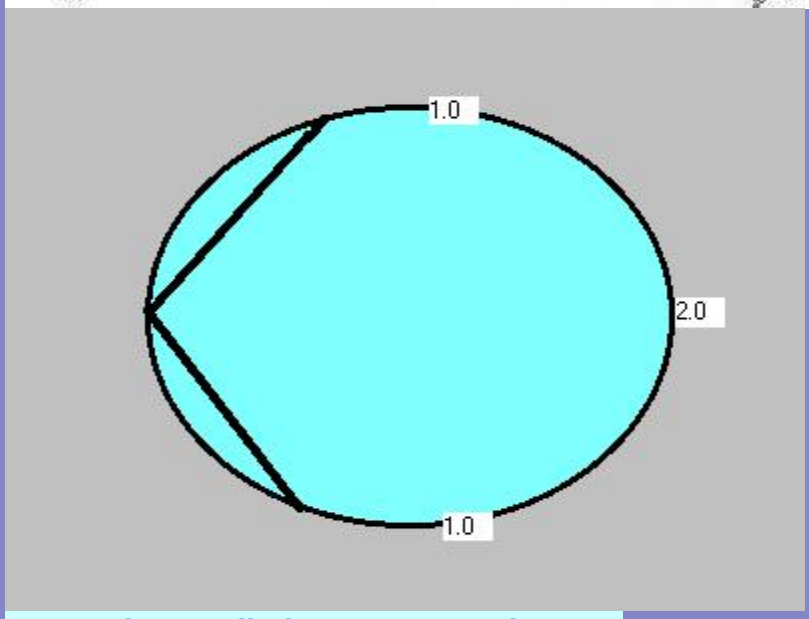
Using the 36 inch & 38 inch extenders together with the whips puts the antenna on 17 Meters, Previous experiments with a different length whip on the 36 inch extenders enabled operation on 12 Meters. I lost this bit of research and suppose that with adjustable extenders (which I haven't designed yet) would do the trick of getting multi-band coverage, but below 20 Meters the antenna would lose the ability of 1 man erection. Remember this antenna is directive and is rotated when used in the field and is done by the arm strong method, and if guyed, use guy wires on a slip ring.

THIS IS THE BEST PERFORMING ANTENNA I HAVE EVER BUILT!
DON'T LET THE SIZE FOOL YOU.
All that is required to use this antenna is a willingness to give it a try and then look at the results. It then comes down to operator skill as to how many contacts you can do per minute.
It truly is AMAZING!
Darrell Koranda KB4XJ

Reference 1 The ARRL Antenna Book 1994, page 13-2 (chart gain of single long wire over a dipole\O-db), page 13-5 (2 long wires placed in Horizontal V, give a 3 db gain over chart on page 13-2)
Reference 2 The ARRL Antenna Book 1994, page 10-20 (forward tilt of elements, gain increase of 4 db) log-periodic V array as compared to log-dipole

Editors note: If you decide to "copy or print" the construction details for this project please note they are in two different text blocks. Copy one at a time for best results. Please give credit to **KB4XJ** for this fantastic project!





Top View Radiation Pattern estimate with field strength meter

Front to back Field strength ratio 18db

Extender is 38" long for 21.250mhz with 6'10 7/8' whip.
Extender is 36" long for 21.350mhz with same whip.
Extender is 38" + 36" for 17 meters with srw 1.4 to 1

Extender element scetch scaled down for detail

3/8 X 24 LONG NUT >>>

3/8 X 24 BOLT 1 1/2 " LONG <<<<<WITH 3/8 X 24 NUT

WELD
WELD

COAX CONNECTION AND MOUNT

BROKEN DOWN FOR FIELD DAY PACKING

TRUCK WHEEL USED AS BASE >>

EXTENDER ELEMENTS

THE ONE ELEMENT V BEAM!

A PROJECT BY KB4XJ

element attaches here>>>

<<<<and here

FOOTNOTES ADDED BY DARREL..... KB4XJ

The dipole formula of 468/freq will work but, an extra 2 inches must be added to the final result.

The induction between the elements changes the resonant frequency of the antenna, and with a dipole cut for 21.250 Mhz (SWR 1:!) when folded into the V shape, the induction of the elements moves the freqency UP to 21.350 Mhz (SWR 1:1).
The Basic principle of the V antenna, when dealing with the rf radiation lobes are that the lobes in the bisector of the V tend to add and the other lobes tend to cancel.

A CHALLENGE

I'am President of the Local Ham Radio Club, and the chairman of the upcoming Field Day 2002. My 20 Meter operator has vowed revenge for the beating he got in last years Field Day from the One Element Beam and has stated that he's going to give me a run for the money and has bought a NEW antenna which will replace his 2 element butterfly beam.
I accepted his challenge and told him I would be using my V and that I wanted his competition AND that I wasn't going to roll over and give up just because he got a new antenna.

The proto-type 6 Meter Horizontal V will be used by Spencer Whitmire W4ERC, the Vice President of the local radio club, Spencer was on 10 Meters with a 2 element beam and was converted to a believer when he saw the V in operation in last year's field day.
Spencer will give 6 Meters a try with the V.
Our callsign for Field Day 2002 will be K4W (Kilo Four Whiskey).

I became an EXTRA in 12-4-98. The One Element Beam is the results of several years of work on various antennas, I was really depressed with the results of tests on so called big signal antennas and finally went back to the horizontal dipole with good reports. I remembered from my beginning in ham radio about the inverted V that I used on 75 Meters and how I could get into Hawaii. I also remembered that an inverted V could be made directive by the leg angle. This was the turning point which led to the ONE ELEMENT BEAM!

While reviewing the logs of last years field day, I found KH6, VE6, VE3, VE4, VE5, KP4, VE7, NP2, DJ2, NP4, TZ6, GO6, VE1, VE9, XE1, RX3, DL6, EA4, OH4, G4, G0, EA5, EA3, DL1, OH1,

RA3, DJ2, F6, PA3, F5, F6, AND VE2. I found these prefixs in the log book of the DX worked during field day, with the Horizontal V. Darrell...KB4XJ

E-Mail address kb4xj@strato.net

"Many thanks to Darrell for all his efforts, hard work and DEDICATION TO HAM RADIO while getting this great project up and running for all hams to enjoy!"
N4UJW HAMUNIVERSE.COM

BACK TO ANTENNA LAB

SCETCH BY JOHN BUTLER
SON OF N4UJW

15822

Why pay full retail for inkjet supplies when you can save between 50%-70%! Click here and start saving!

NEW!
20 THRU 6 METER VERSION
BY
LA0HV OF NORWAY
CLICK HERE!

PLEASE KEEP HAMUNIVERSE FREE
CLICK HERE

Tri- Bander Project Coming Soon from KB4XJ

A comparison of 10 meter verticals using modeling

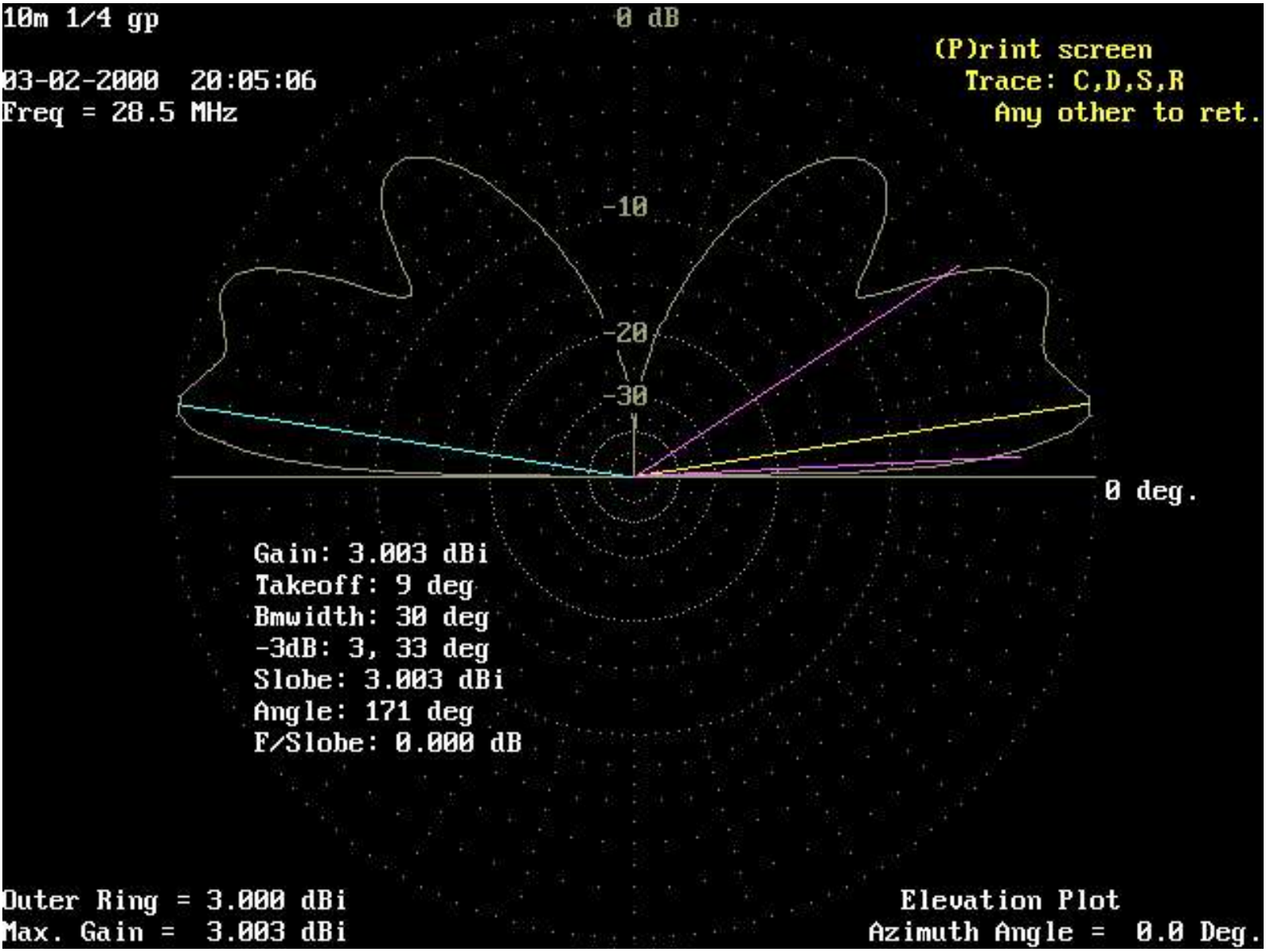
I decided to test a few versions of popular ten meter verticals using elnec demo to test. I have had trouble in the past trying to convince people I was using a 5/8 ground plane with decent success. And I could see their point as normally a 5/8 ground plane has a fairly high angle of maximum radiation. But a major difference was I have been using 3/4 wave radials instead of the common 1/4 wave versions. I found what I think is the reason I have had fairly good success. But I also found a few things I was doing wrong in earlier versions. I found for maximum gain, I would be better off using 5/8 wave radials instead of the 3/4 wave. I had assumed in the past that being the 3/4 wave radial would be resonant and the 5/8 would not be,that the 3/4 would be better. But modeling this antenna, this doesn't seem to be the case. I found using 5/8 radials and sloping at a steep angle to the ground, would give the most gain. With the antenna acting more as a dual 5/8 collinear than a 5/8 radiator with a 5/8 wave ground plane.

Don't have elnec or eznec? You can download a demo version here at http://www.eznec.com/elnecdem.htm

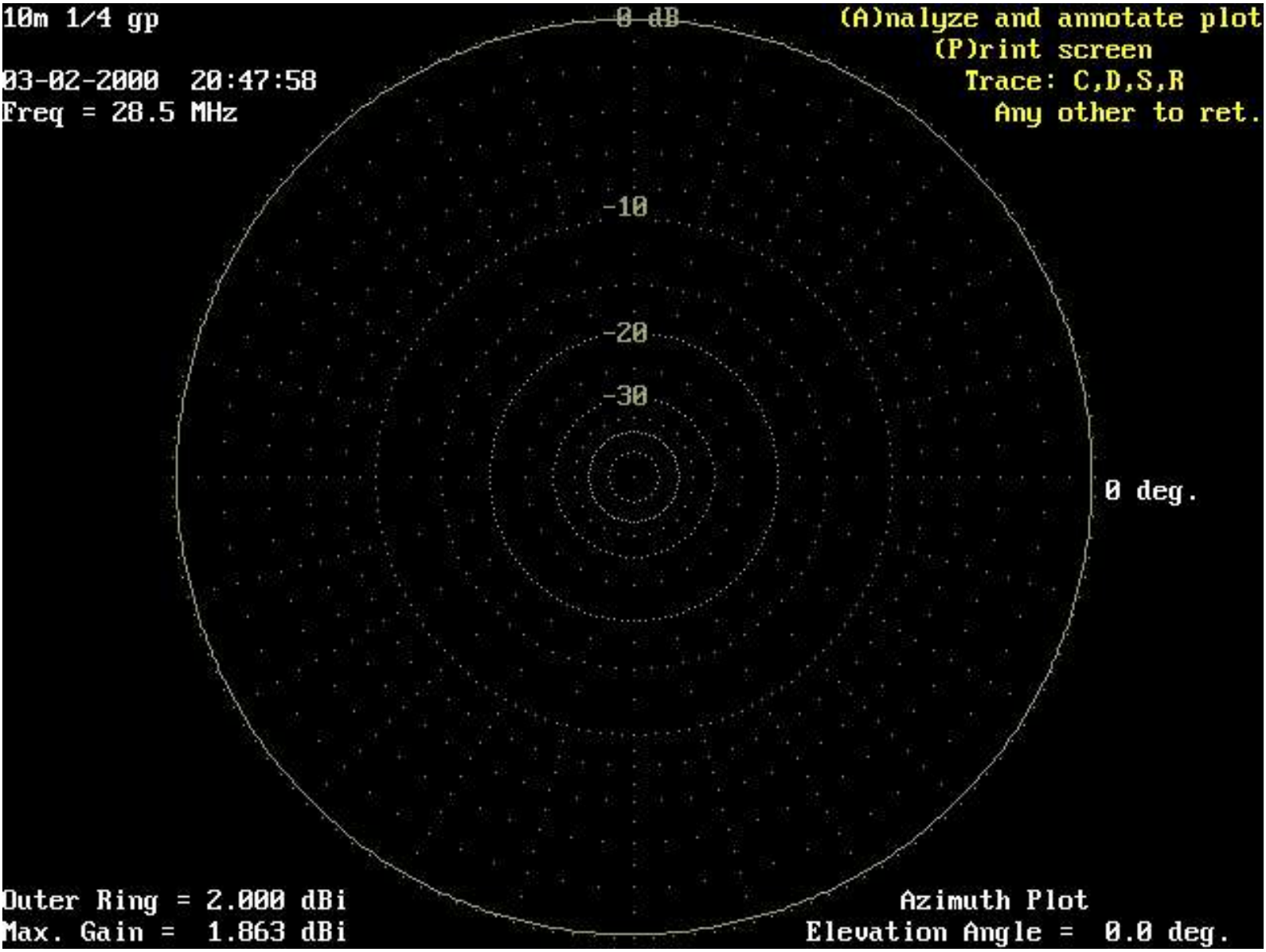
---

I will show the plots of the popular 10 meter verticals. All antennas are modeled at the 40 ft level. I'll start with the 1/4 wave ground plane with 4 sloping radials.

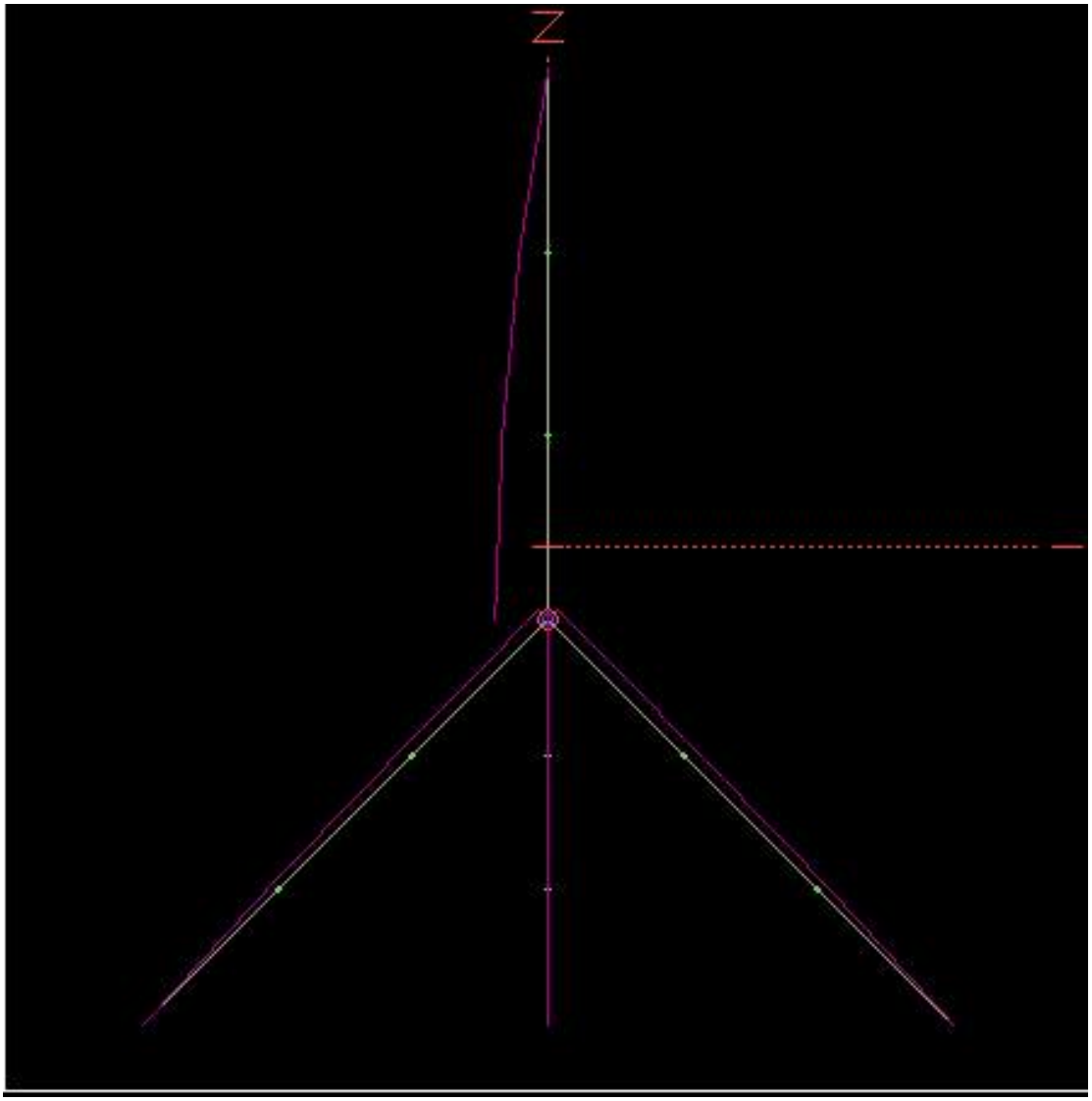
This is the elevation plot over average ground with the antenna at 40 feet.



And the azimuth plot showing freespace gain.



And a view of the antenna showing the current distribution



Page 1

Go to page 2-The half wave vertical

Page

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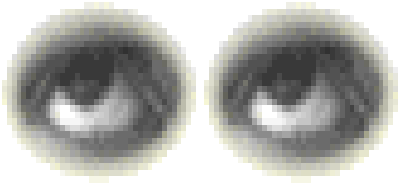


Back to Homepage

Christoph Petermann Homepage

The DF9CY Three-Element Antenna for 28 MHz

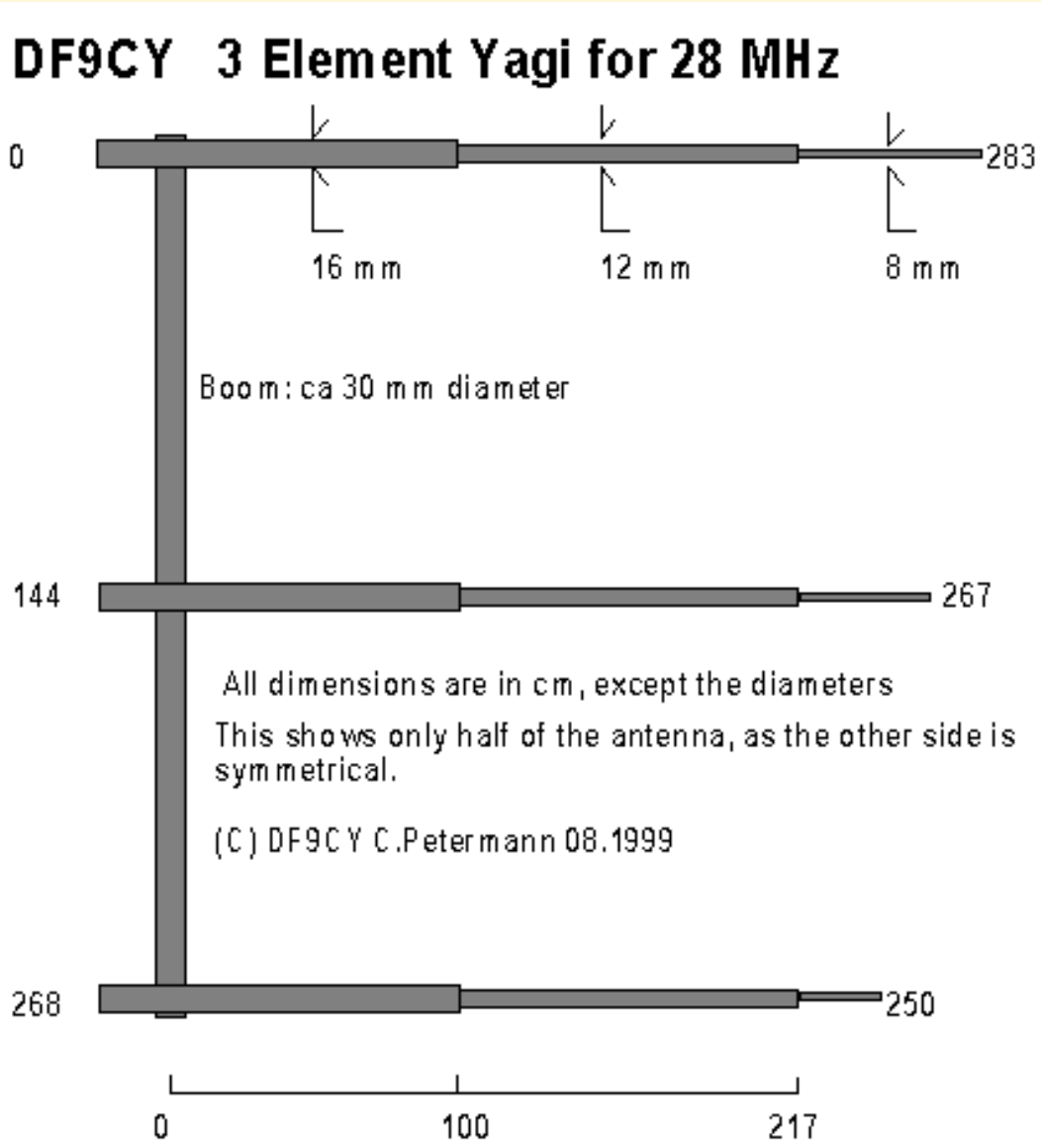
Christoph Petermann DF9CY 1999

A better view of the 3 element for 28 (and the 6 element for 50 MHz)

My 3-Element for 28 MHz antenna is mounted below my 6 element for 50 MHz

The 3 element is a commercial antenna by PAN International, which I slightly modified for good performance on the 10m Amateur-Radio band.
Design Frequency: 28.300 MHz

- **Usable Bandwidth: 600 kHz**
- **Return loss: better than 20 dB with center at 28.270 MHz**
- **Gain: > 7.5 dBi**
- **Front/Back ratio: > 20 dB**



This antenna is a commercial CB Radio antenna distributed by *PAN International*. For this kind of antenna it looked quite robust and indeed it survived a wind storm with wind speeds up to 160 km/h.

Feeding the antenna

This antenna is fed by a gamma match. Setting up this antenna was very easy. Position of the gamma match as approximately moving the bracket as far inside as possible and the inner conductor "looks out" by 70 mm at the end.

Modify the elements

Calculating the antenna with EZNEC gave already very good results. The redesign for 28.270 MHz resulted in shorting each elements' end by 40 mm. This is indeed very easy, as there exist two holes for self-cutting screws separated by 40 mm. Move the outer element in by one hole, fix it with a screw and drill a hole for the second screw. Ready. You shall better not use the screws delivered with the antenna as they tend to oxidation (rust) very

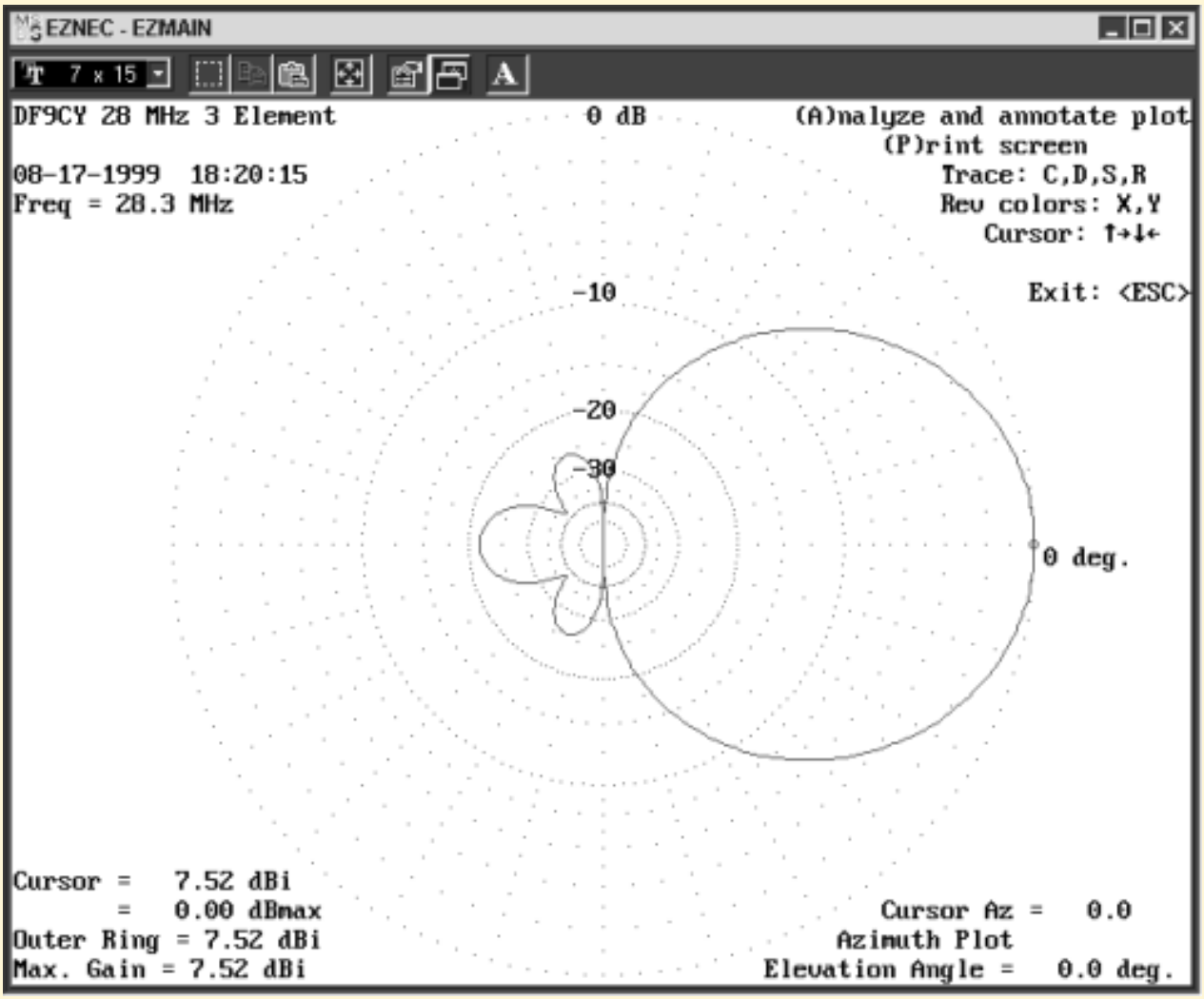
quickly. I bought a complete new V2A set of screws and I have no problems. The elements are electrically connected to the boom. What I dislike is the SO239 connector "popped" to a sheet of metal. I will replace this by a N-Connector now.

Performance

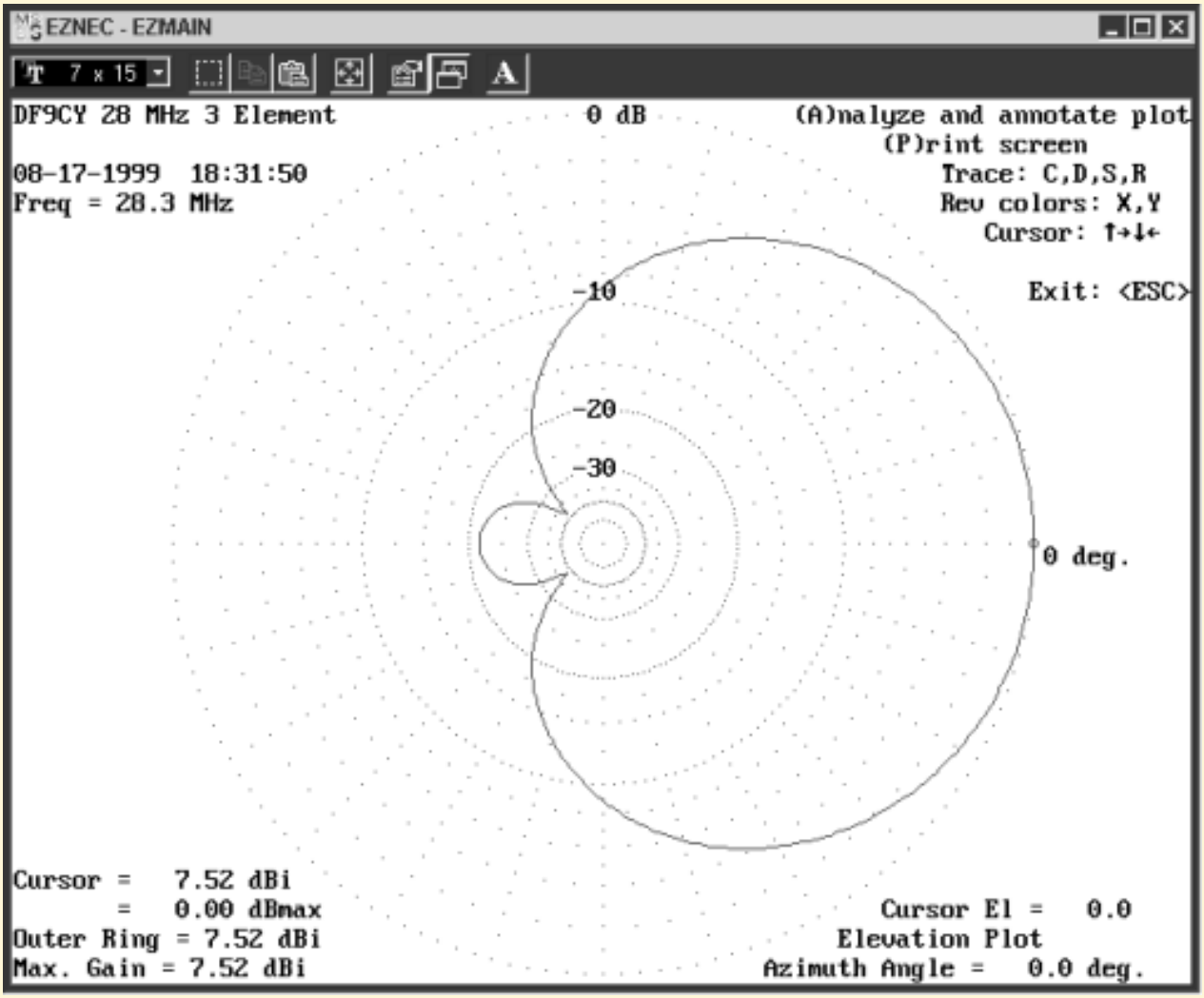
The antenna is up since October 1998 and performance is good. Contacts were made with all continents using my TS690 with about 90 Watts output. It does not perform as our "pile-up-cracker" does, but I feel it works better than a tribander. The antenna is installed at about 9 m above ground. This is indeed not very high, but the surrounding terrain descents in most directions putting the antenna virtually to a higher position. The feedline I use is a low-loss UHF cable (AirCell 2000) with less than 0.4 dB of loss on the 22m run. I also used it with 750 Watts output from a 3-500 for a while and it did a fine job.

EZNEC
Here is the EZNEC simulation file: **cy28-3dopt.zip**

DIAGRAMS AND SIMULATIONS



Azimuth diagram of the 3 Element for 28.3 MHz. Note the good front to back ratio.



Elevation diagram. Vertical attenuation is around 10 dB.

My special thanks go to Roy Lewallen W7EL for programming this fantastic *EZNEC 2.0*

If you want to have more background knowledge on antennas and how-to make them, you will find an ultimate resource here:

W4RNL
L.B.Cebik W4RNL ultimate ANTENNA Site. A must !

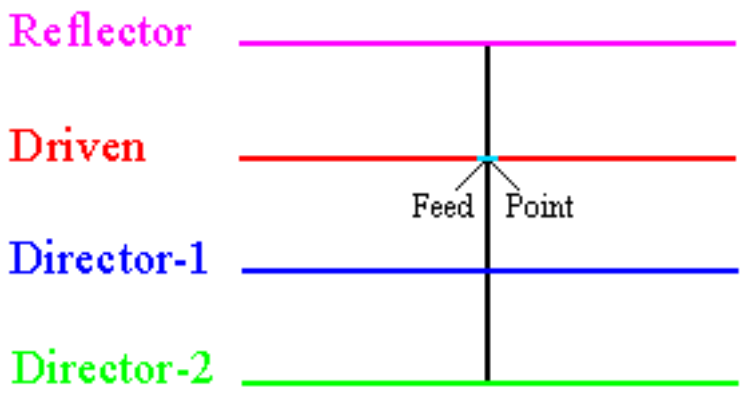
Mail any comments to: df9cy.petermann@t-online.de

Text and All Images are Copyright by Christoph Petermann DF9CY

GO (back) and visit my **homepage**

VHF / UHF Direct Connect Beams

Here are some lengths and spacings for various direct connect beams. The layouts are straight forward and are illustrated below. Your SWR should be less than 1.3:1 with these designs.



The driven element is cut into two halves and insulated from the boom with nonmetalic material. Then the two wires of the coax are connected, one to each section of the driven element. You may drill small holes and use sheet metal screws to accomplish this.

The reflector, director 1, and director 2 can be attached directly to the boom by a variety of methods. There are some helpful hints for antenna construction on our antenna construction tips page.

70 Centimeter 2 element beam center frequency 440 MHz

70 Centimeter 2 element beam 1/8" diameter tubing	Element Length	Element spacing from Reflector
Reflector	12-3/8"	*****
Driven	12-1/8"	6-1/2

	Freespace	Over ground 30 feet
Gain	4.17 dbd	8.08 dbd @ 5 degrees 9.48 dbd @ 14 degrees
F / B	4.01 db	4.0 db

70 Centimeter 4 element beam center frequency 440 MHz

70 Centimeter 4 element beam 1/8" diameter tubing	Element Length	Element spacing from Reflector
Reflector	13"	*****
Driven	12"	8-1/16"
Director 1	11-7/8"	16-3/4"
Director 2	11-3/4"	23-3/8"

	Freespace	Over ground 30 feet
Gain	8.19 dbd	12.78 dbd @ 5 degrees 13.05 dbd @ 14 degrees
F / B	11.15 db	11.2 db

---

2 Meter 4 element quad center frequency 145.000 MHz

2 meter 4 element quad #12 copper wire	Element Length	Element spacing from Reflector
Reflector	21-1/4"	*****
Driven	21"	23-1/8"
Director 1	20-1/4"	46-1/4"
Director 2	20-3/16"	74"

	Freespace	Over ground 30 feet
Gain	8.87 dbd	10.24 dbd @ 5 degrees 13.87 dbd @ 8 degrees
F / B	8.57 db	8.7 db

---

2 Meter 4 element beam center frequency 146.52 MHz

2 Meter 4 element beam 1/2" diameter tubing	Element Length	Element spacing from Reflector
Reflector	38-1/8"	*****

Driven	36"	24-1/4"
Director 1	34-1/2"	49"
Director 2	34-3/8"	71-1/2"

	Freespace	**Over ground 30 feet**
Gain	8.43 dbd	10.34 dbd @ 5 degrees 13.71 dbd @ 10 degrees
F / B	13.32 db	13.60 db at 5 degrees

---

2 and 6 Meter 4 element quad center frequency 146.520 / 52.000 MHz

2 and 6 Meter 4 element quad #12 copper wire	**Element Length**	**Element spacing from Reflector**
6 Meter Reflector	60-1/4"	*****
6 Meter Driven	57-9/16"	48-3/8"
2 Meter Reflector	21-1/4"	54"
2 Meter Driven	20-1/2"	77-1/4"
2 Meter Director 1	20-1/4"	100-1/4"
6 Meter Director 1	57-1/16"	104-3/4"
2 Meter Director 2	20-3/16"	128"
6 Meter Director 2	49-1/2"	181-15/16"

	Freespace	**Over ground 30 feet**
Gain	8.3 / 8.35 dbd	13.97 dbd @ 2 degrees 13.79 dbd @ 8 degrees

F / B	7.86 / 15.84 db	8.12 / 16.59 db

6 Meter 3 element beam center frequency 52.000 MHz

6 Meter 3 element beam 1" diameter tubing	**Element Length**	**Element spacing from Reflector**
Reflector	109"	*****
Driven	101-5/8"	64-15/16"
Director 1	96-3/4"	125-3/16"

	Freespace	**Over ground 30 feet**
Gain	6.5 dbd	9.99 dbd @ 5 degrees 12.01 dbd @ 8 degrees
F / B	8.99 db	9.25 db

6 Meter 4 element beam center frequency 52.000 MHz

6 Meter 4 element beam 1/2" diameter tubing	**Element Length**	**Element spacing from Reflector**
Reflector	109-7/8"	*****
Driven	104-1/4"	72"
Director 1	100.5"	138"
Director 2	100"	204"

	Freespace	Over ground 30 feet
Gain	8.2 dbd	11.78 dbd @ 5 degrees 13.76 dbd @ 8 degrees
F / B	14.44 db	15.28 db

10 Meter beam center frequency 28.450 MHz

10 Meter 3 element beam 1/2" diameter tubing	Element Length	Element spacing from Reflector
Reflector	207-1/2"	*****
Driven	192"	105"
Director 1	185"	210"

	Freespace	Over ground 30 feet
Gain	6.33 dbd	5.49 dbd @ 5 degrees 11.49 dbd @ 10 degrees
F / B	10.21 db	10.37 db

Antenna Construction Tips
Back to the Antenna Elmer
Back to the Home Page

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A 5 BAND LOG PERIODIC DIPOLE ARRAY

In ARRL Antenna Compendium No. 4, I desribed how I modified the Telerana which was orginially featured in the July 1979 issue of QST. The unique feature of the Telerana is that the elements are wire instead of aluminum tubing which makes for a light weight LPDA (log periodic dipole array). The array is suspended within a framework made of fiberglass poles emanating from a central hub with the ends tied together with light weight rope around the perimeter.

Here is a picture taken of the Modified Telerana taken from below.

Modifications to the orginal Telerana design included adding parasitic reflectors for 15 and 20 meters to improve the front to back ratio on these bands. I also described how I added a trap 30/40 meter inverted Vee that acts as a top truss system to keep the Telerana from turning upside down like an umbrella might in high winds. Below the modified Telerana is a homemade 4 element 6 meter Yagi. This system allows coverge of all amateur radio bands from 40 to 6 meters without an antenna tuner, on one tower, with only 3 coax feed lines. A very compact system that works very well.

I have been using this antenna system since 1986. It is supported by a Wilson telescoping tubular tower with the modified Telerana at 62 feet and the apex of the 30/40 trapped inverted Vee at 72 feet. I have worked 293 contires and confirmed QSLs from 275 countries with this system! As well, I have worked 121 countries at QRP levels!

Vol 4 of the ARRL Antenna Compendium series is no longer in print however I have scanned a copy of the article, "The Improved Telerana, with Bonus 30/40 Meter Coverage, here

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You may also find an abbreviated version of the Modified Telerana article in recent ARRL Antenna Handbooks.

Using Moxon Rectangles for WARC-Band Antennas
Part 2: Some 30-17-12 Meter Ideas

L. B. Cebik, W4RNL

Last month we looked at the Moxon rectangle, cut for 17 meters, as the basis for a very compact dual-band Moxon-Yagi for 17 and 12 meters. The design used open-sleeve coupling between the physically driven 17-meter driver and the slaved 12-meter driver. The Yagi portion was a standard driver-director design with about 0.07 wavelength spacing. The result was a two-band array about 10' long and 20' wide that provided over 6 dBi free-space gain, better than 20 dB front-to-back ratio, and a direct feed with 50-Ohm coax (with, as always, a recommended choke or 1:1 balun to attenuate common-mode currents).

Compactness: this goal was one of the good reasons for using a Moxon as the basis for the dual band beam. The width of a Moxon is only about 70% that of a full size Yagi with a driver and reflector, and the Moxon uses no loading to achieve the shortening. Instead, it bends the elements around to point toward each other. By selecting the correct proportions and tip-to-tip spacing, one can obtain a parasitic driver-reflector array with nearly the gain of the full size Yagi and considerably improved front-to-back ratio over the Yagi. Part of the reason for the excellent front-to-back performance lies in the current magnitude and phasing on the rear element relative to the front element: they are close to what one might obtain with each element driven to perfection for maximum rearward rejection.

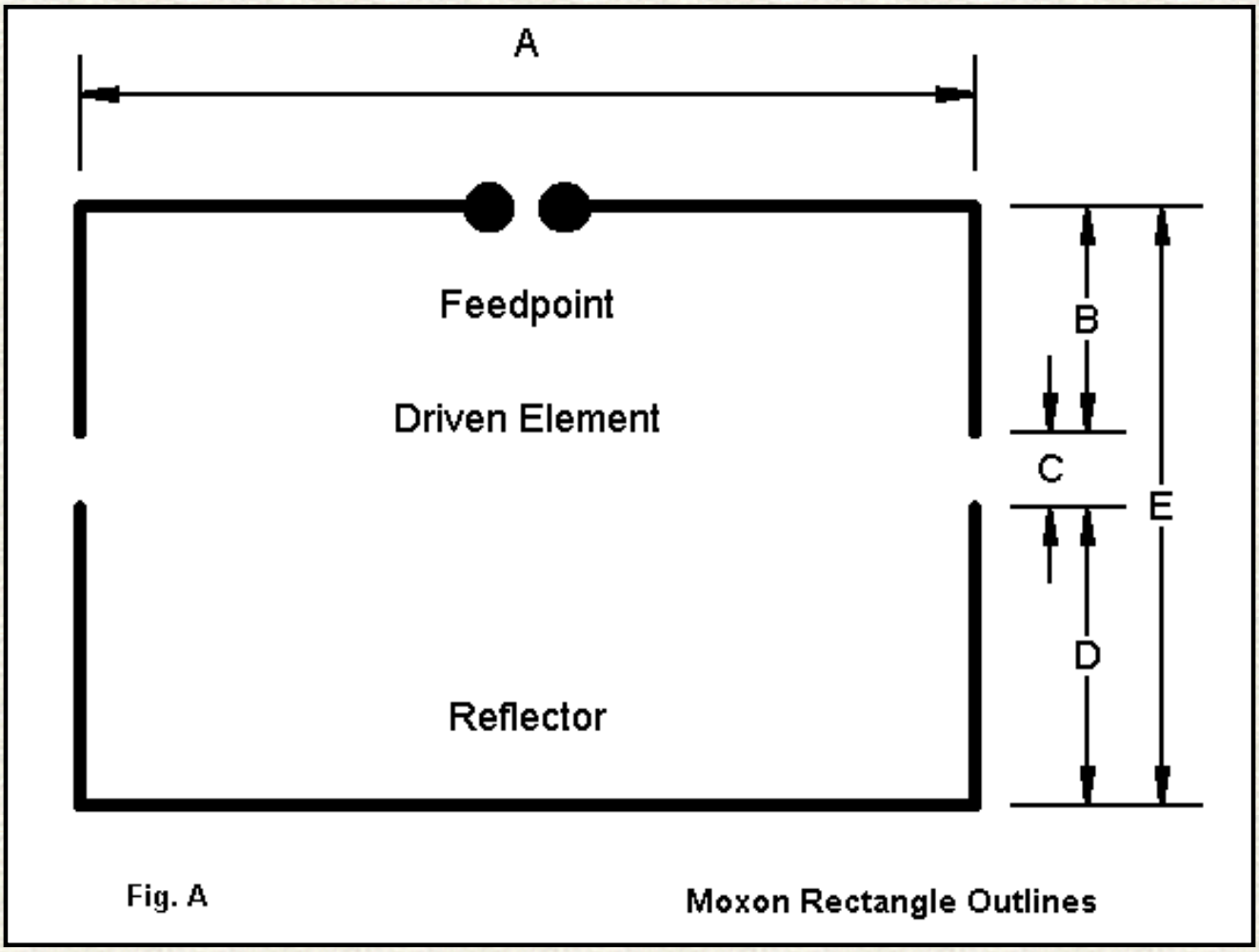


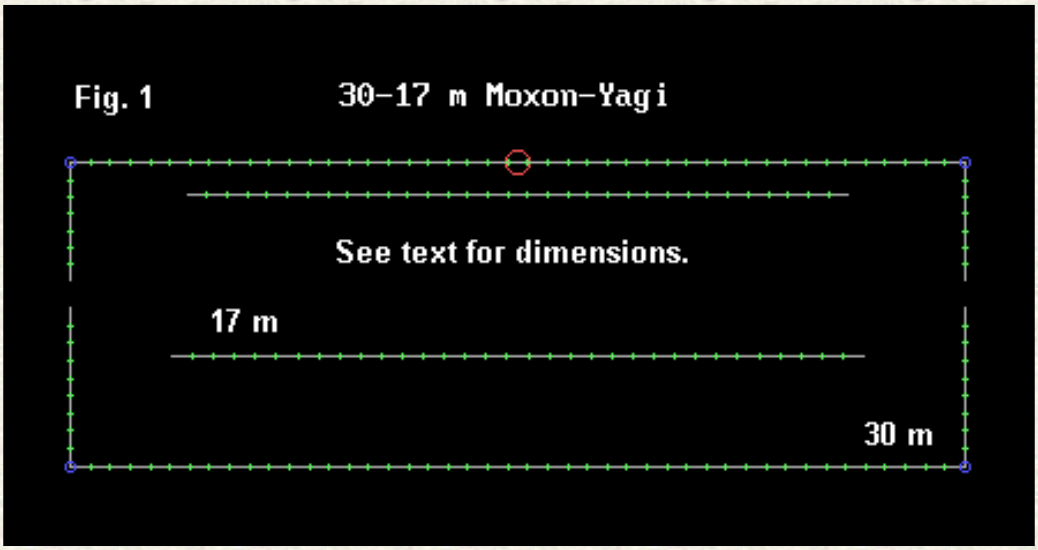
Fig. A provides an outline sketch of a basic Moxon, a refresher. As well, the portions of the Moxon structure are identified, since we shall once more provide some design ideas involving this antenna. However, this month, we shall include 30 meters.

A 30-meter Yagi with unloaded elements would be about 48' to 49' side-to-side. A Moxon rectangle for 30 meters requires only 35' of side-to-side space, and about 13' front-to- back. These dimensions are close to those for a common 20-meter beam. The structure of a 30-meter Moxon might have to be a bit beefier than that of a 20-meter Yagi, since the parallel element must support the "tails" (B and D in **Fig. A**). Nonetheless, for those with space limitations, 35' elements are usually easier to sustain than 48' elements.

A 30-17 m Combination

When we examined the possibilities for 17 and 12 meters, we reached two practical conclusions. First, a dual Moxon array may be too sensitive to minor variations to be truly practical for home building. Second, placing a full size 12-meter Yagi of driver- reflector design inside the 17-meter Moxon was not feasible due to the length of the 12- meter elements--especially the reflector.

Therefore, we shall bypass a dual Moxon for 30 and 17 meters. However, we shall not forego the possibility of a full-size 17-meter Yagi placed within the frame of a 30-meter Moxon. The longest element of a driver-reflector Yagi for 17 meters is 27' and that should be no problem within the 35' dimension of the 30-meter Moxon. **Fig. 1** shows the general outline of the combination.



The following table of dimensions uses the designators of **Fig. A** for the Moxon. The Yagi element spacing entries are distances from the Moxon driver. All dimensions are in feet. The Moxon elements are 1.25" aluminum tubing, while the Yagi elements are 1" diameter aluminum tubing.

Band	Dimension	Length (feet)
30-meter Moxon	(all elements use	1.25" diameter aluminum tubing)
Moxon	A	34.91'
	B	4.89'
	C	1.12'
	D	6.56'
	E	12.57'
17-meter Yagi	(all elements use	1.0" diameter aluminum tubing)
Driver	Length	25.80'
	Space	1.30'
Reflector	Length	27.00'
	Space	8.00'

The spacing between the Yagi elements is 6.7' or about 1/8 wavelength. The spacing of the Moxon elements is about 0.13 wavelength. The Moxon dimensions are unchanged from those optimized for maximum front-to-back and a 50-Ohm feed when used independently. Of course, changes in material dimensions or the use of stepped-diameter elements will require readjustment of the design to yield satisfactory performance. As I did last month, I shall place model descriptions at the end of these notes for those who wish to experiment with other material combinations.

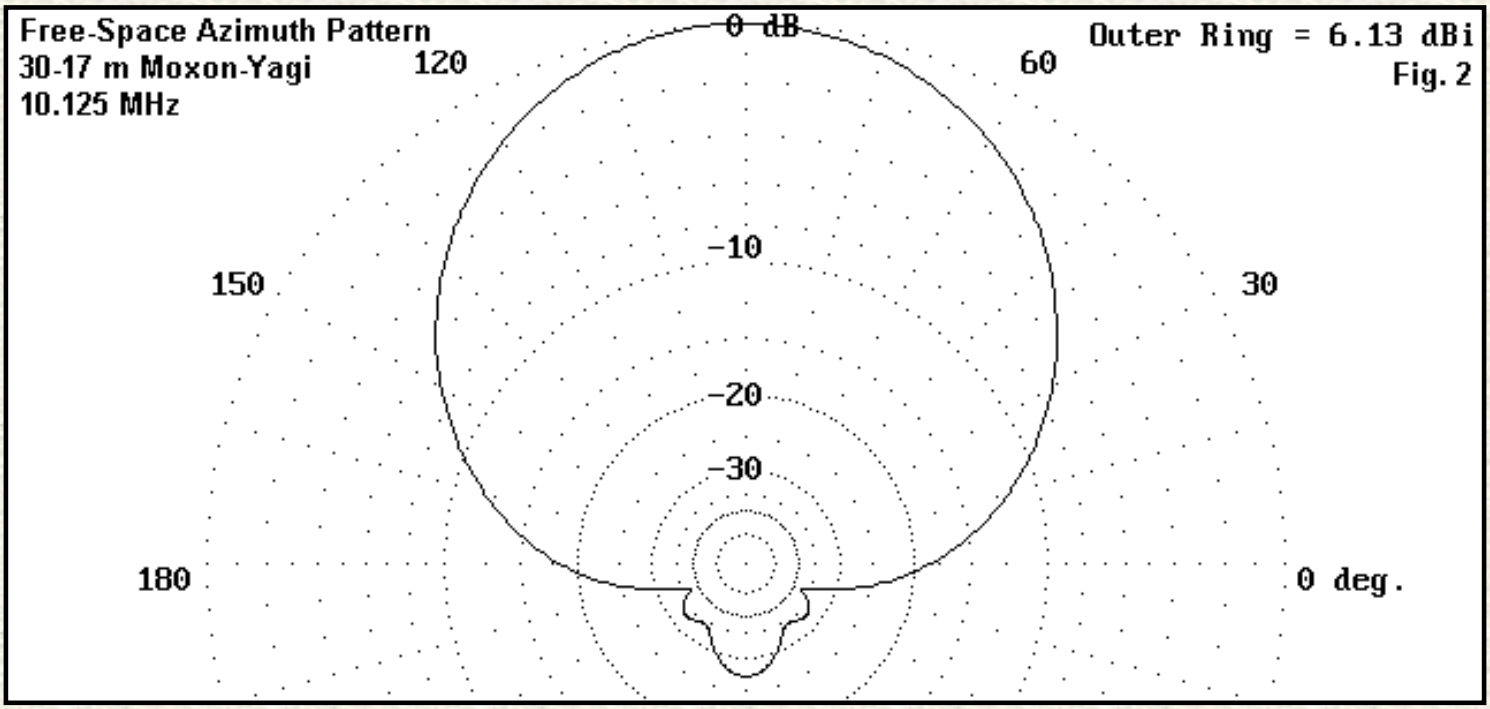
However, once the basics are established, the only post-construction adjustment will be to the slaved 17-meter driver. Its exact length and spacing from the Moxon driver will determine the feedpoint impedance on 17 meters at the physical feedpoint.

The Moxon-Yagi combination is a very well-behaved. The following performance table, listing the gain in terms of free-space gain, gives a good general picture.

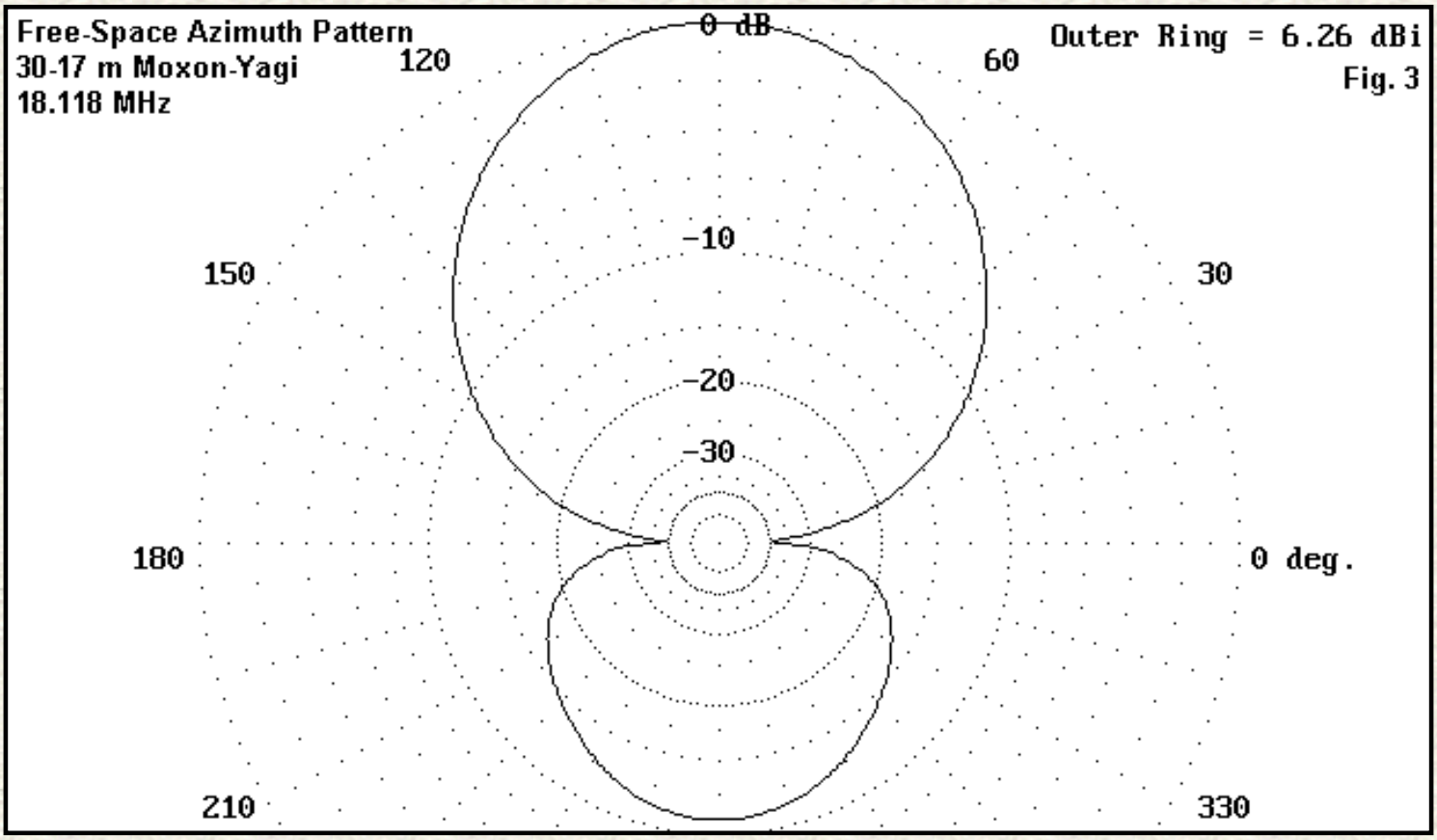
Frequency MHz	Gain dBi	Front-to-Back Ratio dB	Feedpoint Z R+/-jX Ohms	50-Ohm VSWR
10.100	6.21	22.9	47.6 - j 3.6	1.09

10.125	6.13	27.0	50.7 - j 1.4	1.03
10.150	6.05	33.4	53.7 + j 0.6	1.07
18.068	6.37	10.8	52.4 - j22.5	1.55
18.118	6.26	10.9	49.9 - j 5.4	1.11
18.168	6.16	10.9	47.5 + j10.9	1.26

As we saw last month, the addition of a closely-spaced parasitic beam slightly detunes the Moxon in terms of moving the peak front-to-back ratio upward in frequency. However, the gain and the source impedance are mostly unaffected. **Fig. 2** provides a free-space azimuth pattern of the 30-meter Moxon performance at mid-band. One of the side-benefits of the Moxon rectangle design is that it tends to hold its high front-to-back ratio even down to a height of 3/8 wavelength--about 36' on 30 meters. Of course, the old rule that higher is better still applies. Nevertheless, even a modest installation can expect quite reasonable results on 30 meters.



On 17 meters, we must expect the lesser front-to-back ratio associated with the driver- reflector Yagi design. **Fig. 3** shows the mid-band free-space azimuth pattern for the array at 18.118 MHz.

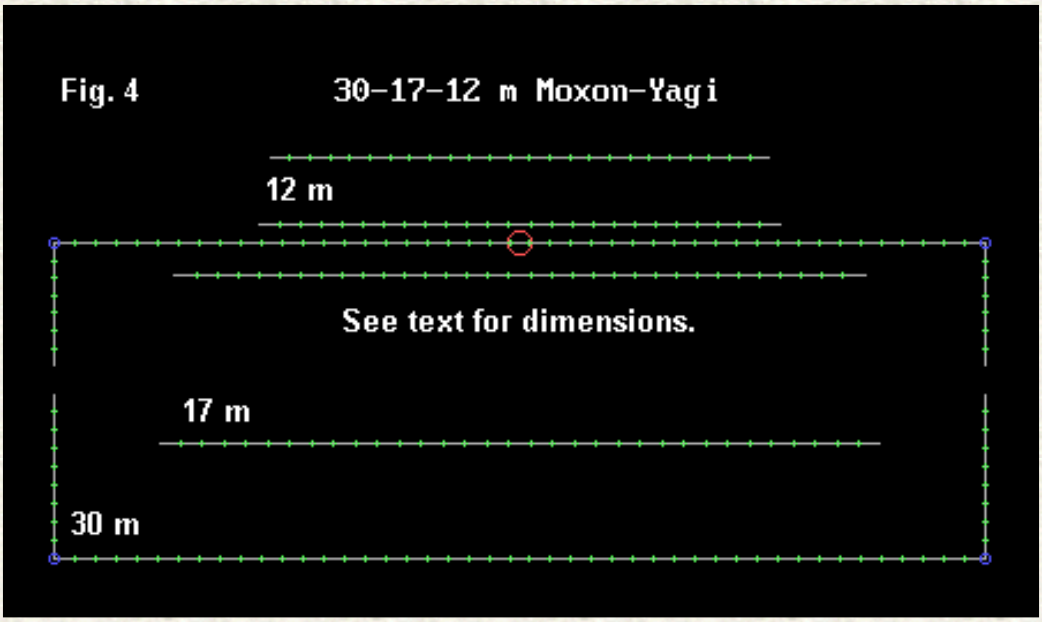


Because the upper band Yagi is behind and within the Moxon rectangle, there is no significant forward-stagger effect. Hence, the Yagi gain and front-to-back ratios are virtually identical to those one might obtain from an independent driver-reflector Yagi for 17 meters. However, with 1/8 wavelength spacing, the feedpoint impedance of an independent Yagi would be closer to 35 Ohms. The near-50-Ohm match is obtained by virtue of the open-sleeve coupling, which can be set for virtually any desired impedance by changes in the length and/or spacing of the driver relative to the physically fed element. As we saw last month, the impedance changes more rapidly on a slaved element than on a directly fed element. In the case of the 17-meter Yagi, it is the reactance that undergoes the most rapid change, while the resistance remain quite stable.

Although the 17-meter Yagi does not gain anything from being inside the Moxon, it does not lose anything either. Moreover, it does not take up an additional space on a supporting tower. The two-band array has identically the same outside dimensions as the 30-meter Moxon itself. Except for the additional weight and wind load of the 17-meter elements, there seems little reason not to add the upper band if one seriously plans a 30-meter Moxon.

A 30-17-12 m Combination

Suppose we might be willing to extend the boom of the Moxon-Yagi from 12.6' to about 16 feet. For the additional 3.4' of boom length, we add one more band to the array--with virtually no change in the design work done so far. A driver-director Yagi with an element spacing of about 0.07 wavelength can be added ahead of the Moxon driver to arrive at a WARC 3-band array of considerable compactness. **Fig. 4** shows the general outline of the arrangement.



For the lower bands, the design once more uses 1.25" diameter elements for 30 and 1" diameter elements for 17. The 12-meter elements are 0.5" in diameter, and everything is aluminum. With these materials we can uses the dimensions in the table below. For the 17-meter elements, the spacing entry indicates the distance behind the Moxon driver. For 12 meters, the spacing entry indicates the element spacing forward from the Moxon driver.

Band	Dimension	Length (feet)
30-meter Moxon	(all elements use	1.25" diameter aluminum tubing)
Moxon	A	34.91'
	B	4.89'
	C	1.12'
	D	6.56'
	E	12.57'
17-meter Yagi	(all elements use	1.0" diameter aluminum tubing)
Driver	Length	25.80'
	Space	1.30'
Reflector	Length	27.00'
	Space	8.00'
12-meter Yagi	(all elements use	0.5" diameter aluminum tubing)
Driver	Length	19.46'
	Space	0.70'
Director	Length	18.70'
	Space	3.43'

For the Moxon and the 17-meter Yagi, nothing has changed. The dimensions of the 12-meter Yagi are very slightly different from those used with the 17-meter Moxon last month. However, that Moxon used elements with a smaller diameter than the 30-meter Moxon in this design. The major change is in the spacing from the Moxon to the 12-meter slaved driver--somewhat wider than in the 12-17 design.

If we see little difference in dimensions, we should also expect little difference in performance. The following performance table provides the numbers.

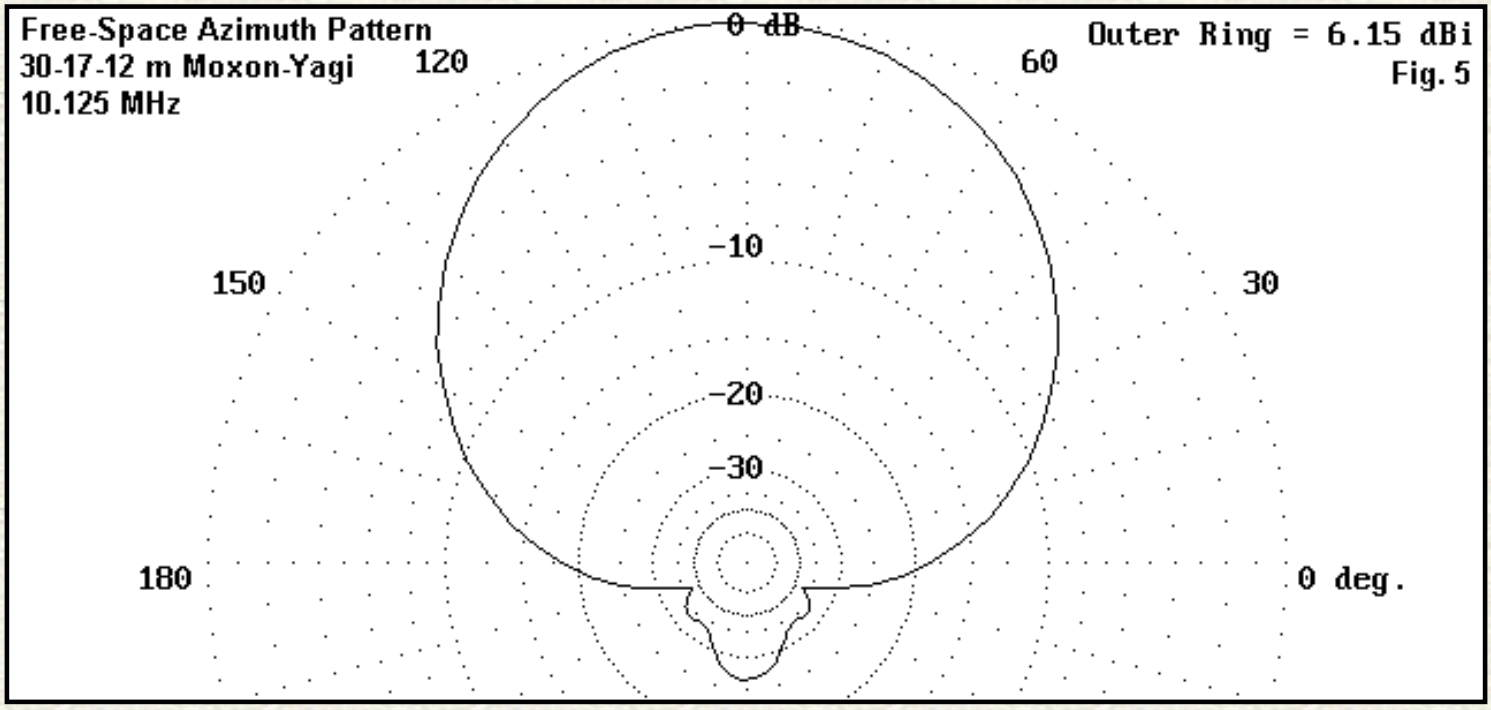
Frequency	Gain	Front-to-Back	Feedpoint Z	50-Ohm

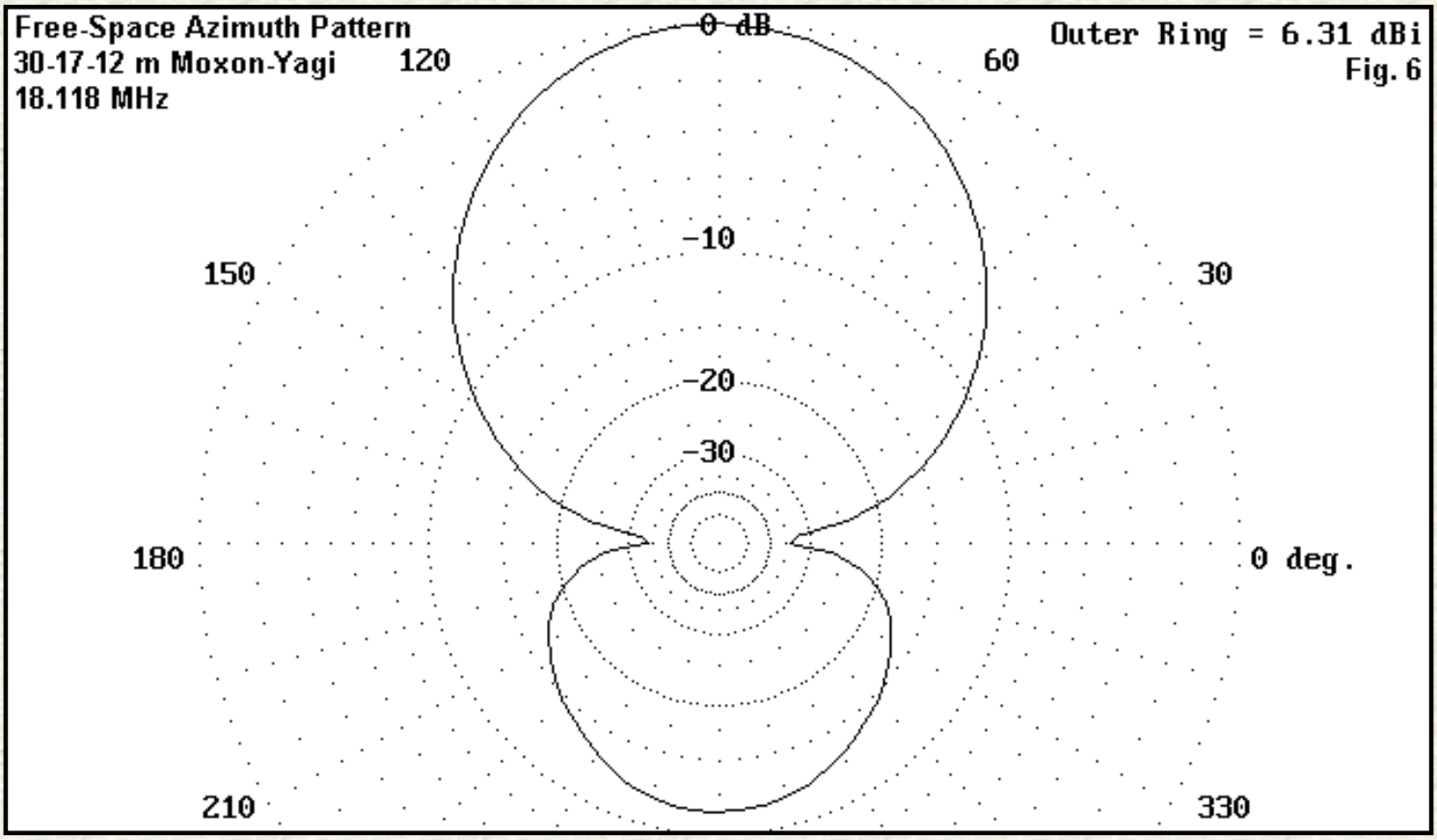
MHz	dBi	Ratio dB	R+/-jX Ohms	VSWR
10.100	6.23	22.5	48.5 - j 4.9	1.11
10.125	6.15	26.3	51.7 - j 2.8	1.07
10.150	6.07	32.1	54.9 + j 0.9	1.10
18.068	6.40	11.3	53.3 - j16.3	1.38
18.118	6.31	11.4	51.3 - j 1.2	1.04
18.168	6.21	11.4	49.4 + j13.6	1.31
24.89	6.57	25.3	59.4 - j 9.9	1.28
24.94	6.67	22.1	48.7 + j 3.4	1.08
24.99	6.76	19.2	38.9 + j18.5	1.62

On 30 meters, performance only shows insignificant changes in the last decimal places of the performance figures. On 17 meters, the performance values are numerically up, but again, not in a way that makes a difference that one could detect in use. The numeric increase is due in part to the forward stagger effect that gives the 12-meter elements a slight director effect during 17-meter operation.

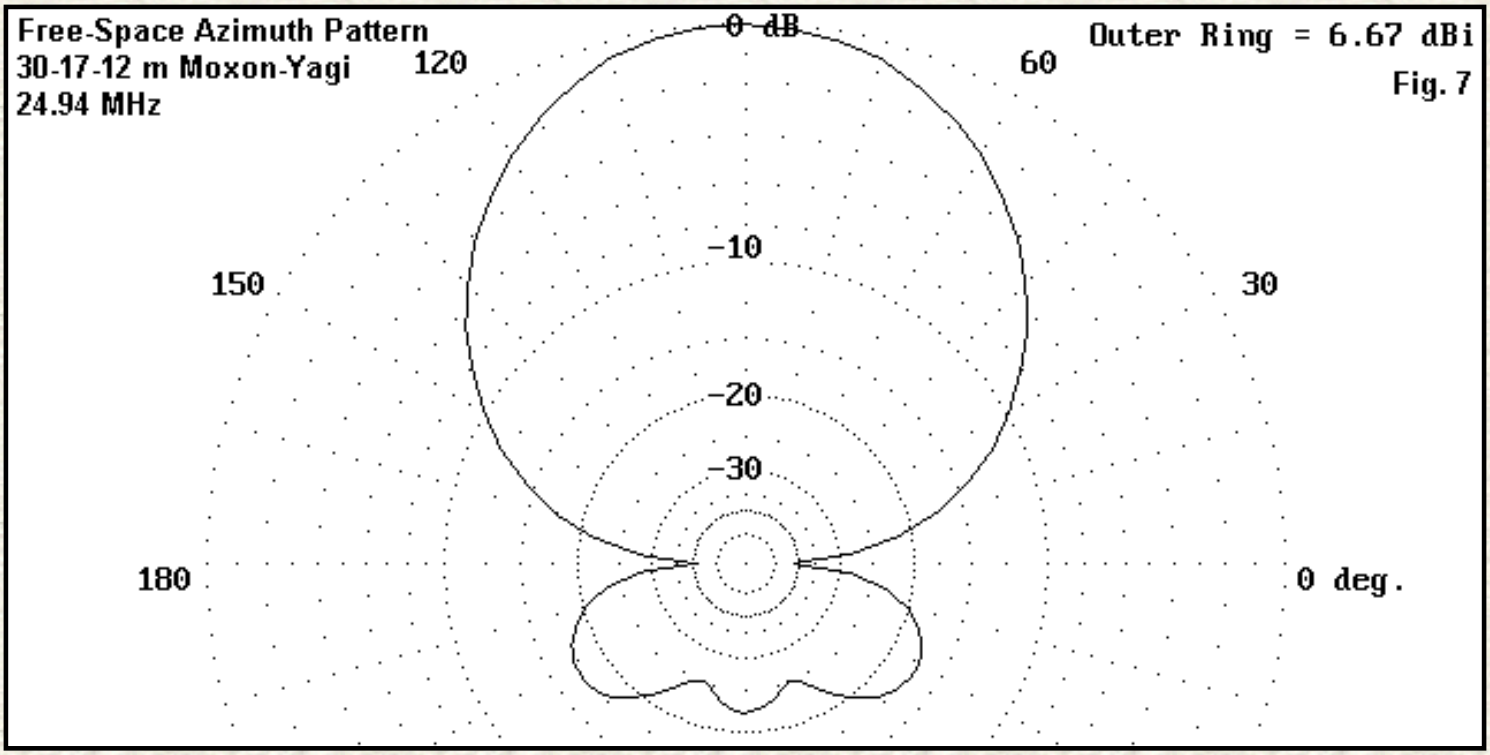
On 12 meters, the 2-element driver-director array performs normally for an antenna of this type, although the independent feedpoint impedance of about 20 Ohms is overcome by slaving the driver to the 30-meter Moxon driver. Last month we saw higher gains on 12 meters. However, the 30-meter driver is farther removed from the 12-meter elements and provides some isolation of them from the interior 17-meter elements. Hence, forward stagger effects are minimal.

Nonetheless, for a beam with a 16' boom length, the tri-band Moxon-Yagi offers excellent potential. **Fig. 5** shows the free-space azimuth pattern at 10.125 MHz.





In **Fig. 6**, we see essentially the same pattern as in **Fig. 3** for the middle of 17 meters. The 12-meter free-space azimuth pattern for 24.94 MHz (**Fig. 7**) is similar to the one shown for 12 meters last month. The same rear quartering lobes are present to moderate the 180-degree front-to-back ratio.



The exact spacing and length of the slaved drivers will require field adjustment to arrive at a final setting to achieve a

good 50-Ohm match on each band. However, those adjustments are about the only critical matters related to construction--once you decide on and design for the precise set of materials you will use. Essentially, you can design an independent Moxon for 30 meters and adjust it to perfection. Then, you can add the 17-meter elements and perform driver adjustments until you are satisfied. Rechecking the 30-meter impedance should show no significant change. Finally, add the 12-meter elements and adjust its driver for a good match on that band. Because element spacing is a bit more critical for driver-director arrays than for driver-reflector Yagis, recheck the element spacing for 12 meters before concluding the adjustment procedure. Once more, the impedances for 30 meters and 12 meters should not have significantly changed by adding the 12-meter elements.

I have good luck in adjusting Moxons and simple Yagis close to the ground by pointing the array as close to straight up as possible, with the reflector anywhere from 4 to 12 feet off tghe ground. The adjustments have held true at heights from 20 to 35 feet up. This technique is not applicable for all beams, but it is worth a try for those who do not relish adjusting beam elements from the top of a tower.

For those who wish to do further design work on models of the two antennas, here are model descriptions that should ease wire coordinate entries for most NEC modeling software.

30-17 m Moxon-Yagi Frequency = 10.125/18.118 MHz.

Wire Loss: Aluminum -- Resistivity = 4E-08 ohm-m, Rel. Perm. = 1

--------------- WIRES ---------------

Wire	Conn.	--- End 1 (x,y,z : ft)	Conn.	--- End 2 (x,y,z : ft)	Dia(in)	Segs
1		-17.457, -4.888, 0.000	W2E1	-17.457, 0.000, 0.000	1.25E+00	7
2	W1E2	-17.457, 0.000, 0.000	W3E1	17.457, 0.000, 0.000	1.25E+00	45
3	W2E2	17.457, 0.000, 0.000		17.457, -4.888, 0.000	1.25E+00	7
4		-17.457, -6.005, 0.000	W5E1	-17.457,-12.569, 0.000	1.25E+00	9
5	W4E2	-17.457,-12.569, 0.000	W6E1	17.457,-12.569, 0.000	1.25E+00	45
6	W5E2	17.457,-12.569, 0.000		17.457, -6.005, 0.000	1.25E+00	9
7		-12.900, -1.300, 0.000		12.900, -1.300, 0.000	1.00E+00	33
8		-13.500, -8.000, 0.000		13.500, -8.000, 0.000	1.00E+00	33

------------- SOURCES -------------

Source	Wire Seg.	Wire #/Pct From End 1 Actual	(Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	23	2 / 50.00	(2 / 50.00)	1.000	0.000	V

Ground type is Free Space

30-17-12 m Moxon-Yagi Frequency = 10.125/18.118/24.94 MHz.

Wire Loss: Aluminum -- Resistivity = 4E-08 ohm-m, Rel. Perm. = 1

--------------- WIRES ---------------

Wire	Conn.	--- End 1 (x,y,z : ft)	Conn.	--- End 2 (x,y,z : ft)	Dia(in)	Segs

```
1          -17.457, -4.888,  0.000  W2E1 -17.457,  0.000,  0.000 1.25E+00   7
2     W1E2 -17.457,  0.000,  0.000  W3E1  17.457,  0.000,  0.000 1.25E+00  45
3     W2E2  17.457,  0.000,  0.000        17.457, -4.888,  0.000 1.25E+00   7
4          -17.457, -6.005,  0.000  W5E1 -17.457,-12.569,  0.000 1.25E+00   9
5     W4E2 -17.457,-12.569,  0.000  W6E1  17.457,-12.569,  0.000 1.25E+00  45
6     W5E2  17.457,-12.569,  0.000        17.457, -6.005,  0.000 1.25E+00   9
7          -12.900, -1.300,  0.000        12.900, -1.300,  0.000 1.00E+00  33
8          -13.500, -8.000,  0.000        13.500, -8.000,  0.000 1.00E+00  33
9           -9.730,  0.700,  0.000         9.730,  0.700,  0.000 5.00E-01  25
10          -9.350,  3.430,  0.000         9.350,  3.430,  0.000 5.00E-01  25

              ------------- SOURCES -------------

Source    Wire      Wire #/Pct From End 1    Ampl.(V, A)  Phase(Deg.)  Type
          Seg.     Actual      (Specified)

1          23     2 / 50.00   (  2 / 50.00)      1.000        0.000       V

Ground type is Free Space
```

The 30-17 or the 30-17-12 combination arrays offer some interesting potentials for small WARC band beams. These design notes aim to whet your appetite for further and improved designs. However, you will have to go some to improve performance on all bands while shrinking the size of these arrays. However, if you do manage the trick, I'd be first in line to find out how.

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The All-Band Center-Fed Inverted-L

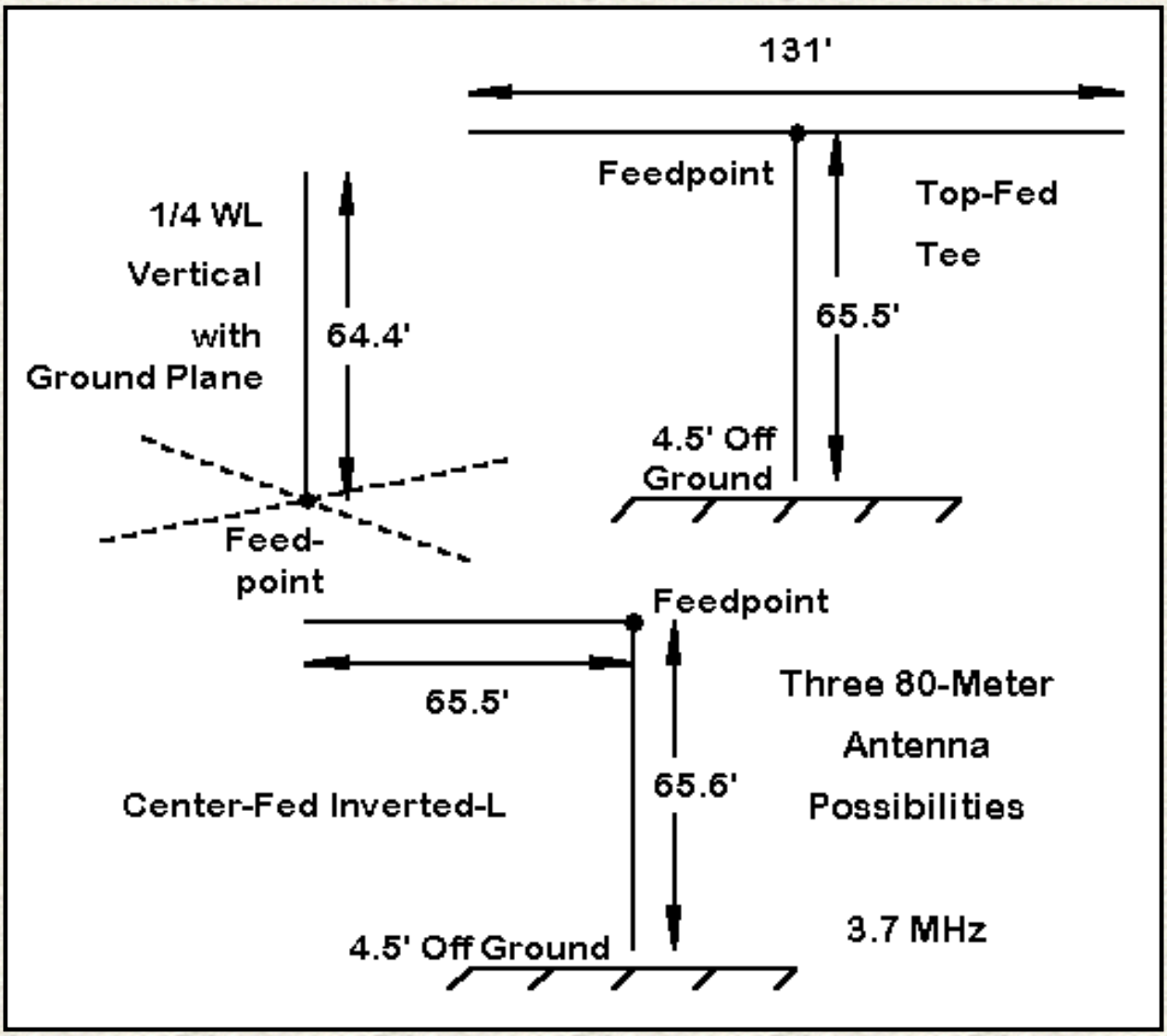
L. B. Cebik, W4RNL

When I wrote on "The L-Antenna" for 10-meters a few months back, I noted that the antenna was not likely new. I have since learned that the basic idea seems to have originated with VK3AM in the early 1950s and is described in L. A. Moxon's (G6XN) classic ***HF Antennas for All Locations*** (pp. 154-156 of the first edition). This antenna is a standing L, although Moxon has no problems with viewing it inverted. Ralph Holland, VK1BRH, includes the L in his computer study of several antennas, including an interesting variant of the L: the 1/2 wl inverted-L. (VK1BRH's interesting modeling studies, published in ***Amateur Radio***, the journal of the Australian Wireless Association, can be found at his web site: http://www2.dynamite.com.au/vk1brh/Antsim.htm) Perhaps the earliest article on the inverted-L as an all-band antenna may have been "The 'Inverted L' Ham Antenna," by Bob See, W5LTD, which appeared in *Radio and TV News*, January, 1959, pp. 64-65. Bob used base feeding to operate the antenna as a standard inverted-L monopole with a ground plane on 80 and as an end- or voltage-fed longer wire above 80, as his measured impedance figures attest. The 1/2 wl inverted-L can also be center-fed using parallel feedline and an ATU. We shall focus on the center-fed version: it is an antenna with excellent potential as an all-band substitute for the 135' center-fed doublet.

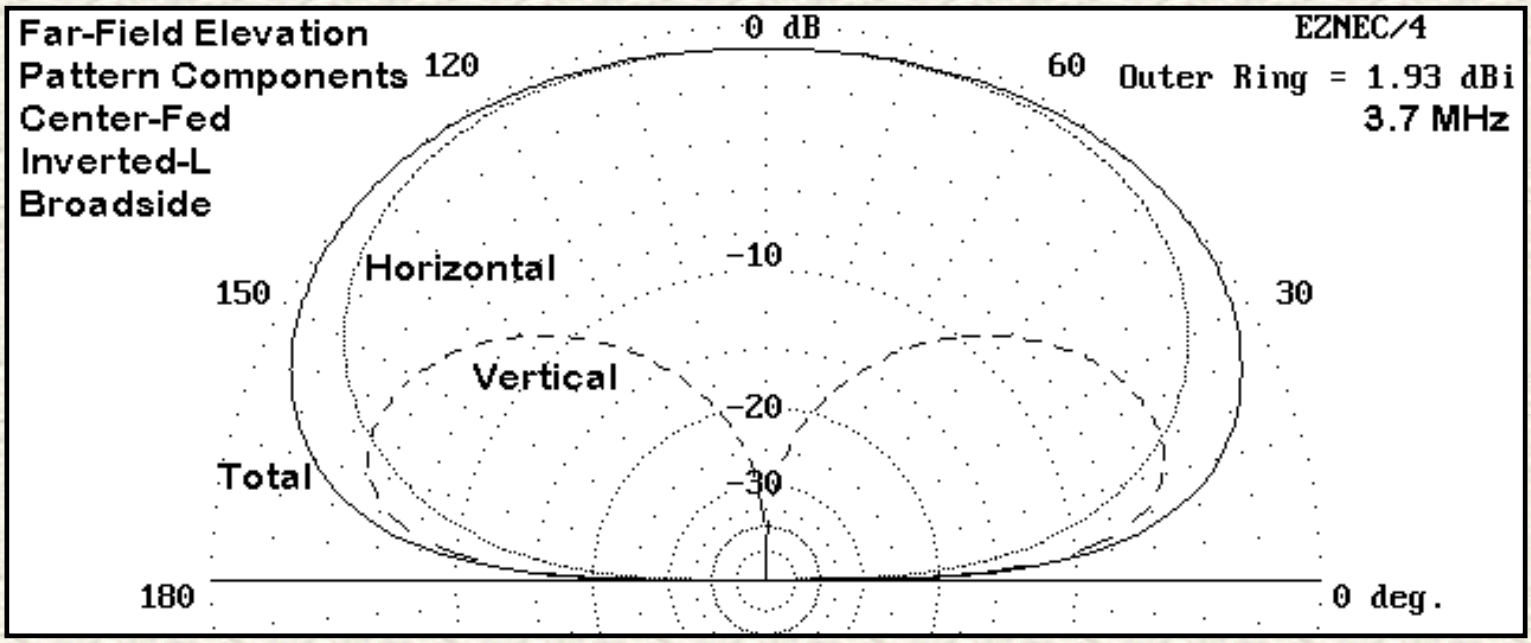
The 1/2 wl inverted-L which we shall examine differs from standard 1/4 wl inverted-Ls in 2 ways: First, it is longer, of course. Second, it is normally current fed at the center (although end- or voltage-feeding is always possible, even if not always convenient). Hence, it can be viewed as an inverted Vee tilted over by 45 degrees. Alternatively, it can be viewed as a 1-leg-ground-plane 1/4 wl vertical upside down.

If the upside-down vertical had a second leg going in exactly the opposite direction, the result would be--to a large degree--cancellation of the horizontally polarized radiation. Let's call this antenna the T.

Both the L and the T differ from the standard 1/4 wl ground plane vertical by being complete antennas--dipoles as it were. Hence, neither requires a ground plane beneath them. For some situations, this fact can simplify construction. The figure below shows the structural differences among the three antennas for models set at 3.7 MHz. The L and T models were set at a top height of 70', with the vertical arm terminated 4.5' off the ground.

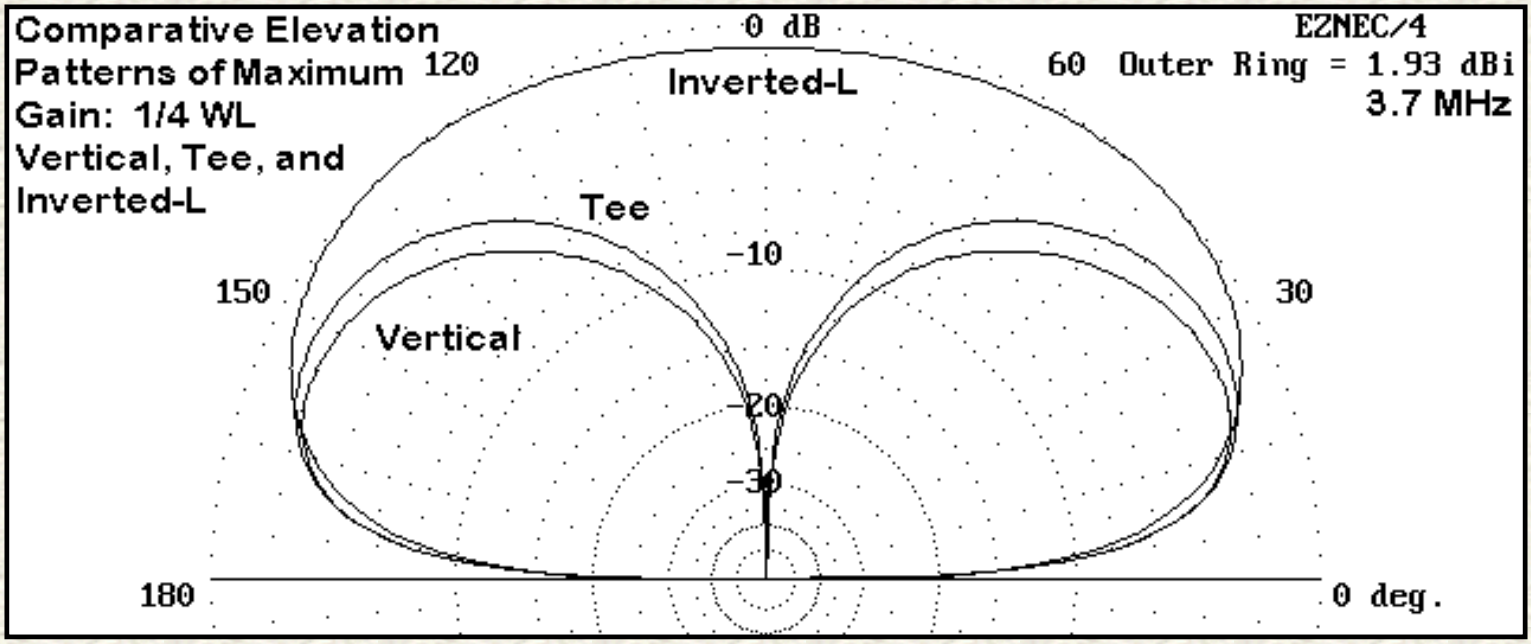


Each antenna was modeled using #14 copper wire and average ground throughout. Note that the inverted-L and the T present challenges to the builder in terms of routing the parallel feedline to the top feedpoint. We shall do some comparisons, but first, let's become a bit more familiar with the inverted-L basic pattern.

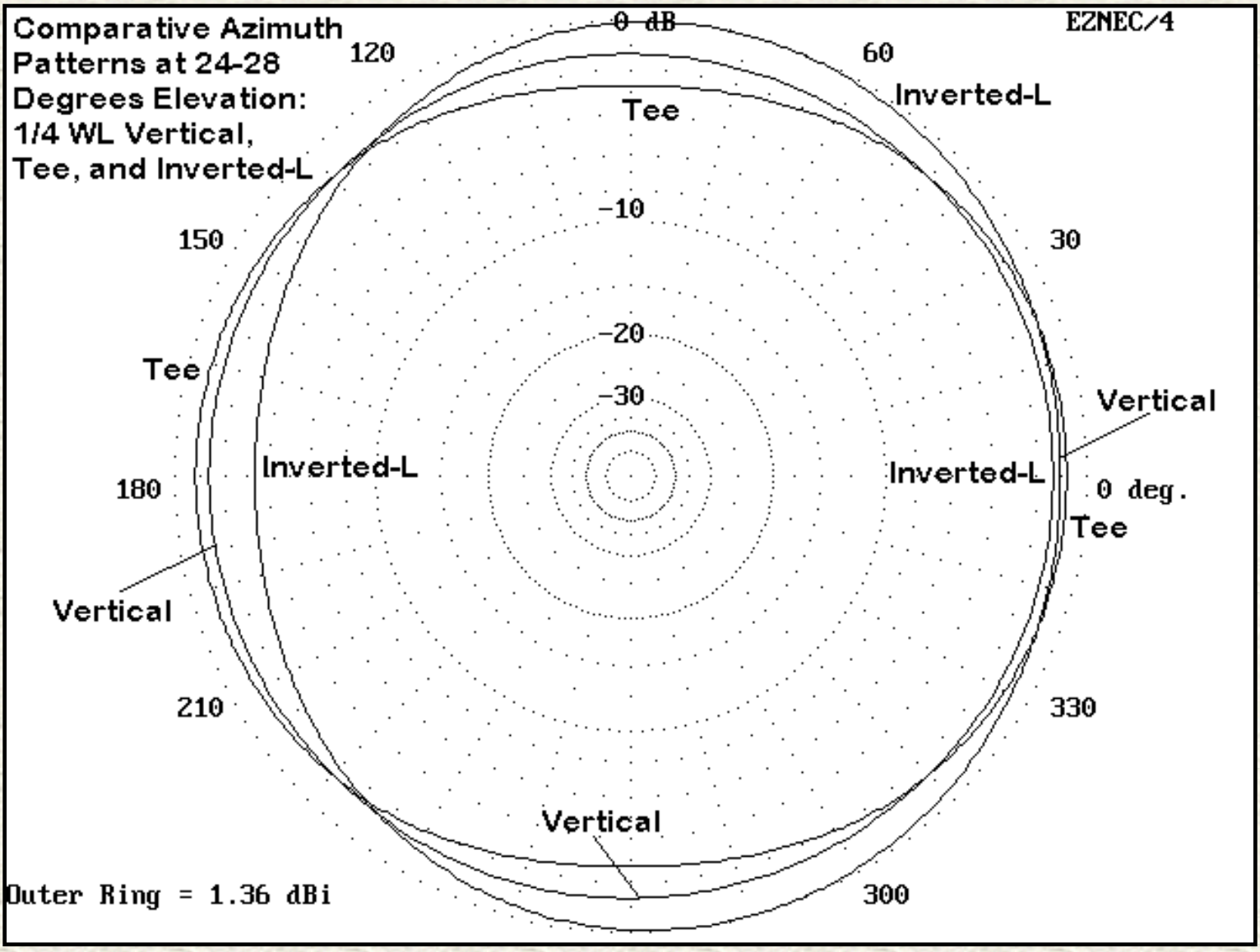


The elevation pattern above shows the vertical, horizontal, and total field components of the inverted-L radiation pattern taken broadside to the horizontal arm of the antenna, where radiation is strongest. In the plane off the ends of the horizontal arm, horizontally polarized radiation is somewhat weaker, but the vertically polarized radiation remains at full strength, with some pattern bending away from the horizontal arm.

A fair comparison might be made among elevation patterns for the L, T, and vertical. Since the total pattern of the L is a broad oval, let's take the strongest direction also of the T, which happens to be off the ends of the horizontal arms. The vertical is truly omnidirectional, so let's set at least 20 radials beneath it.



The comparative pattern above shows the rough equality of the T and the ground-plane-vertical patterns under the specified conditions. Surprisingly, the inverted-L comes close to both antennas in low angle radiation. It also has stronger high angle radiation--without becoming a cloud burner--which is useful for shorter skip contacts. In other words, the inverted-L has potential as an all-purpose low-band antenna.



The azimuth patterns of the three antennas--taken at elevation angles between 24 and 28 degrees--show the slight oval of the T and the slightly more radical oval of the inverted-L. The L's azimuth pattern also shows the slight displacement in the direction away from the horizontal arm. However, these effects are small enough not to stand in the way of using the antenna for general operating purposes.

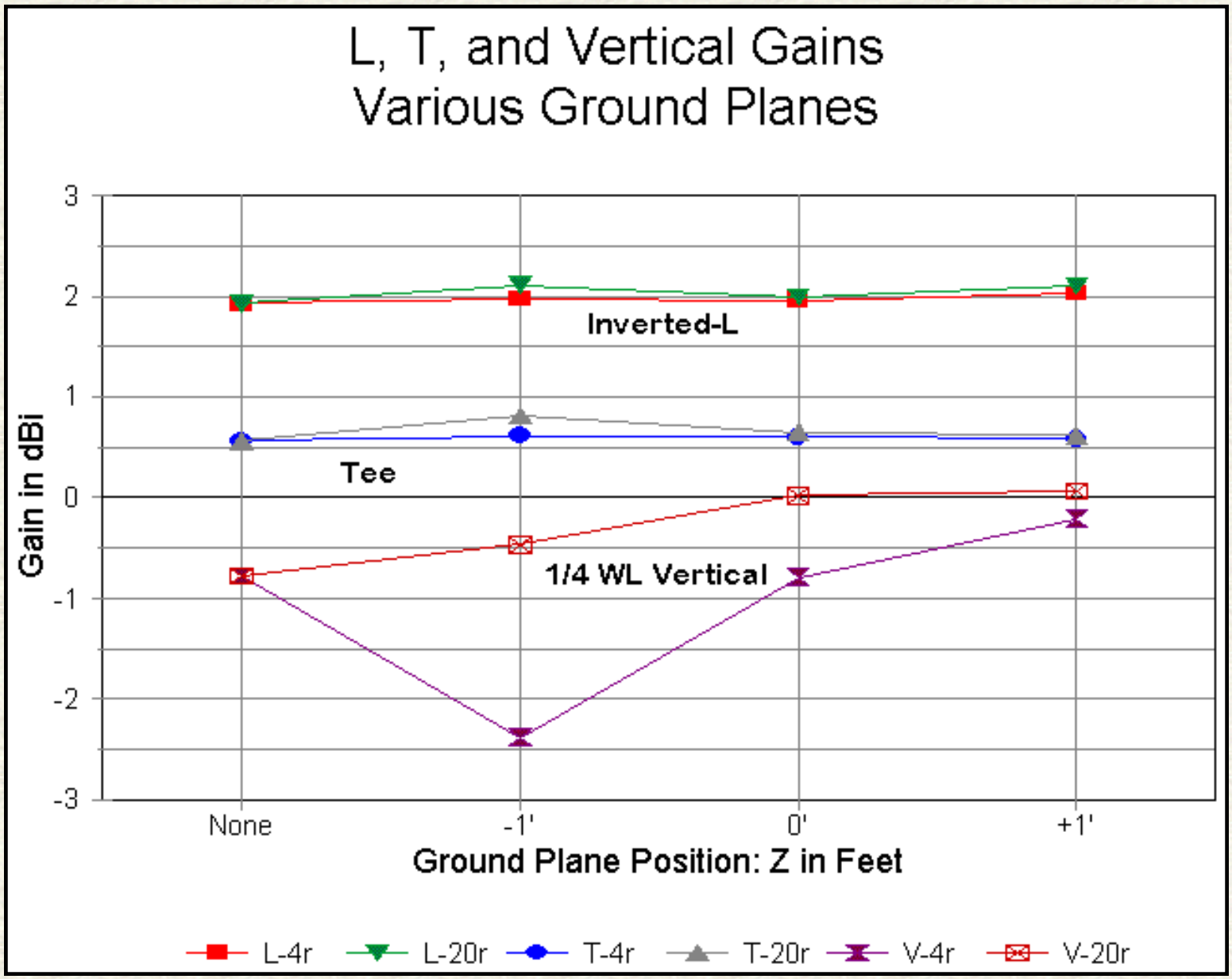
The Ground-Plane Question

In principle, as a complete 1/2 wl antenna, the inverted-L requires no ground plane. Likewise, the T should require none. In contrast, the 1/4 wl vertical requires a ground plane to complete the antenna. To test the relative need and utility of a ground plane, I modeled all three antennas with ground planes, first using 4 wires and then using 20 wires. I set each ground plane first at 1' below ground, then at the surface, and finally at 1' above ground. The vertical's source segment touches the ground, which gives erroneous results in NEC-4. Therefore, the surface ground plane for the vertical was set 0.1' above ground. The radial wires were the same length as the vertical radiators, which means slightly shorter radials for the vertical than for the L or T.

The following table summarizes results for the three antennas with 4 and 20 wire ground planes.

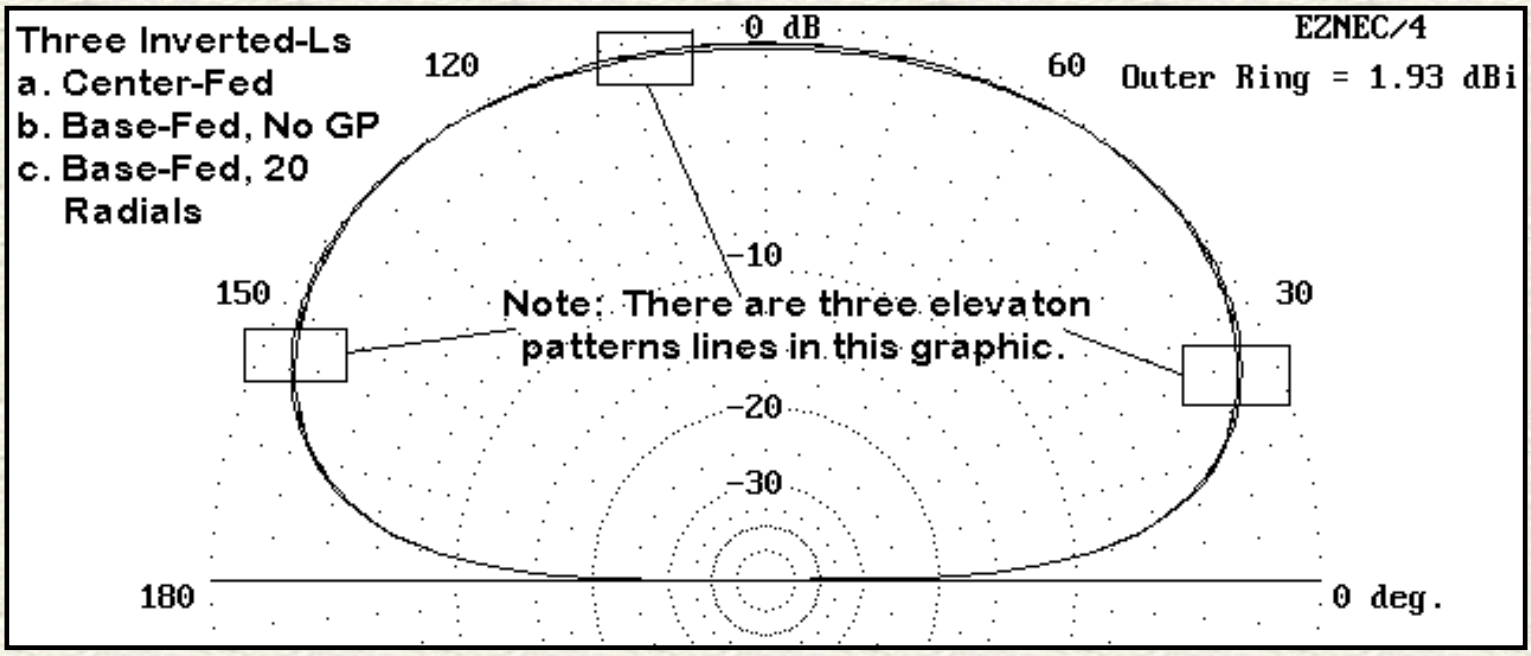
Antenna/GP level	Gain dBi	TO angle degrees	Source Impedance R +/- jX Ohms
4-radial tests			
Inverted-L			
No GP	1.93	44	66.1 + j 3.8
GP -1'	1.97	45	66.1 + j 4.5
GP 0'	1.96	44	66.0 + j 4.0
GP +1'	2.03	46	65.6 + j 4.5
T			
No GP	0.56	28	42.7 - j 5.2
GP -1'	0.61	28	42.7 - j 4.6
GP 0'	0.60	28	42.7 - j 5.0
GP +1'	0.58	28	42.3 - j 5.2
Vertical			
No GP	-0.78	24	48.2 + j 0.2
GP -1'	-2.38	25	68.5 + j 8.1
GP 0'	-0.79	25	45.5 + j32.5
GP +1'	-0.21	24	39.6 - j 9.3
20-radial tests			
Inverted-L			
No GP	1.93	44	66.1 + j 3.8
GP -1'	2.11	45	65.4 + j 6.5
GP 0'	1.99	44	66.3 + j 4.4
GP +1'	2.10	46	64.5 + j 4.7
T			
No GP	0.56	28	42.7 - j 5.2
GP -1'	0.81	28	41.9 - j 3.0
GP 0'	0.65	29	42.9 - j 4.7
GP +1'	0.62	28	41.3 - j 5.2
Vertical			
No GP	-0.78	24	48.2 + j 0.2
GP -1'	-0.47	25	45.3 - j 0.9
GP 0'	0.02	24	37.1 - j 3.9
GP +1'	0.06	24	36.8 - j14.6

Although the tables give the most data, comparisons are more difficult than with a graph.



The antenna gains are compared in the graph above. The line connections between points are not real connections, but only let the eye tell which data points go together. As is evident, NEC-4 modeling strongly suggests that the addition of a ground plane adds virtually nothing to antenna performance for the inverted-L and the T, both of which we have described as complete antennas. In contrast, the vertical is dependent upon the most extensive (up to 60-100 radials) that a builder can install. (The vertical antenna data point for "No Ground Plane" should be used for reference and does not represent accurate data relative to a real antenna.) I further modeled the vertical with 64 radials. At a depth of 1', the antenna gain increased to 0.19 dBi, while setting them 1' above ground yielded a gain of 0.02 dBi. Modeling has consistently suggested that for perfectly symmetrical ground planes above ground, more than 6-8 radials may be superfluous. This conclusion does not necessarily apply to ground planes that are not perfectly symmetrical.

An alternative to high-altitude center feeding of the inverted-L is to base feed it at the low end of the vertical. Models of this mode of feeding the antenna show patterns quite consistent with those for center feeding, with a source impedance in the neighborhood of 5000 Ohms. Once more, the addition of a ground plane does not aid antenna performance in any way, as the following elevation plot shows.



However, the absence of need for a ground plane should not be mistaken for an absence of a need for a good RF ground. In turn, we should not presume that the ground rod near the shack, which provides AC and DC power grounding for safety, also provides an adequate RF ground. Army tests established a couple of decades ago that a good RF ground needs periodic short (<2') rods connected by a perimeter wire or strap that essentially surrounds the entire station location.

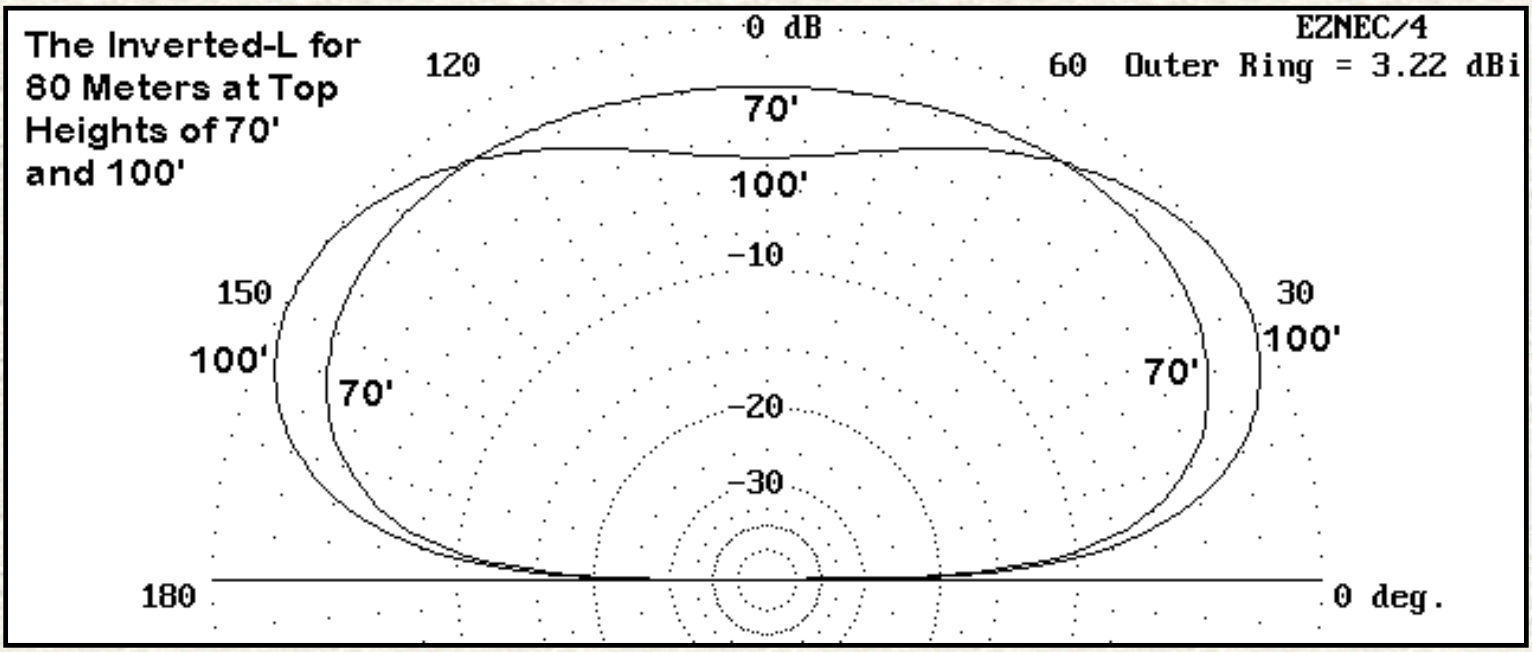
Installation Variations

Knowing that not everyone tempted to use the inverted-L will have all of the space needed, I checked some variants that represent typical construction compromises or changes. Since the antenna will be fed with parallel transmission line, matching is not a major problem. However, changes of gain and elevation angle may indicate that some variations are better than others.

1. Height: elevating the inverted-L is a route to slightly more gain and a lower take-off angle broadside to the horizontal arm. Here is a table of values modeled with top heights at every 5' from 70 to 100 feet up.

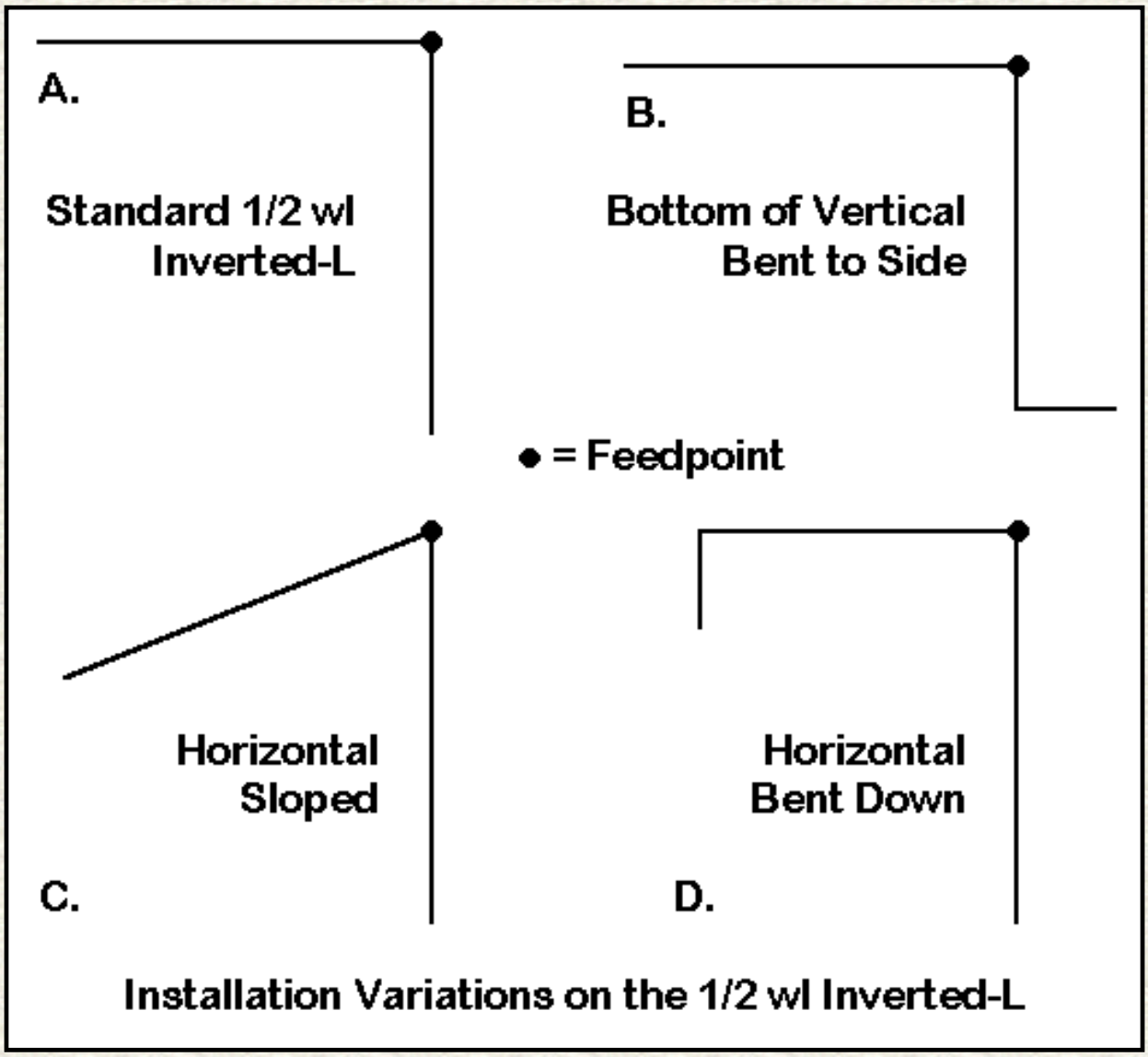
Top Ht. feet	Bottom Ht. feet	Gain dBi	TO Angle degrees	Feed Impedance R +/- jX Ohms
70	4.5	1.93	44	66.1 + j 3.8
85	9.5	2.20	42	62.8 - j 4.2
80	14.5	2.43	41	60.1 - j 9.3
85	19.5	2.63	38	57.6 - j12.8
90	24.5	2.83	37	55.3 - j15.4
95	29.5	3.02	34	53.0 - j17.2
100	34.5	3.22	33	50.9 - j18.5

Nothing drastic happens between any two levels, but the trends are clear. Gain increases and take-off angle decreases. The antenna plays shorter, the higher we go. However, unless one plans to use a monoband coax feed system, the precise dimensions are not at all critical.



The elevation patterns in the figure compare the antenna at 70' and at 100' and add visual confirmation of the conclusion drawn from the table.

2. Sloping and Bending: The more normal problem for home installation is too little vertical or horizontal space. As the figure below shows, there are a number of installation "tricks" we might use. The question at hand is how much each will hurt performance.



Installation Variations on the 1/2 wl Inverted-L

Bending the vertical at the bottom: The first way to save vertical space or to protect family members from the high voltage at the antenna element end is to bend the lower end of the vertical to the side. The upper horizontal arm remains 65.5' long. The overall length of the vertical is also 65.5', but part is now vertical and part horizontal. I tested three scenarios, listed in the table below:

Max. Ht feet	Vert. Wire feet	Low Hor. feet	Gain dBi	TO Angle degrees	Feed Impedance R +/- jX Ohms
70	60	5.5	2.04	45	64.2 - j 4.4
70	55	10.5	2.22	47	61.8 - j11.3
65	55	10.5	1.92	49	64.8 - j 2.7

The chief effect of the bend is to raise the high angle radiation strength a small bit and to raise the elevation angle of maximum radiation. The latter figure indicates a slight loss in the lowest angle radiation, which one would anticipate from shortening the vertical length. None of these small changes in dimension affect the usableness of the antenna.

Sloping the horizontal arm down: One might wish to use the antenna where there is only one truly tall support and the support for the far end of the horizontal arm is lower. The result is a sloping horizontal arm. Using a peak height of 70' and keeping the dimensions of each wire at 65.5', I tested 2 scenarios, representing two degrees of slope, against the

standard installation.

Max. Vert Ht feet	Hor. End Ht feet	Gain dBi	TO Angle degrees	Feed Impedance R +/- jX Ohms
70	70 (no slope)	1.93	44	66.1 + j 3.8
70	60	2.26	51	58.6 + j 8.4
70	50	2.66	58	52.4 + j32.0

Gain increases are at high angles of radiation, with some loss of low angle radiation strength. Although a true horizontal is perhaps the best compromise for maximum low and high angle performance, the patterns with a modest slope to the horizontal arm do not make the antenna unusable by any means.

Bending the horizontal arm far end down: If horizontal space is limited, a common practice is to bend (or dangle) the outer ends of a dipole downward. since the region is the high voltage and low current portion of the antenna, the radiation pattern is least affected by modifying the geometry. Again, I compared 2 scenarios to the full length horizontal arm configuration.

Max. Ht feet	Hor. Arm Lth feet	Bent Length feet	Gain dBi	TO Angle degrees	Feed Impedance R +/- jX Ohms
70	65.5	0.0	1.93	44	66.1 + j 3.8
70	55.5	10.0	1.82	45	62.7 - j 5.9
70	45.5	20.0	1.62	43	54.6 - j12.2

Low angle radiation remains essentially constant, since the vertical arm has not been altered. Further shortening of the horizontal arm would show a gradual further reduction in maximum gain and in the take-off angle. Higher-angle radiation is decreased, although the antenna remains eminently usable.

Like many wire antennas, the inverted-L will tolerate moderate alterations of geometry to fit the space available and still yield good, if not peak, performance.

Multi-Band Use of the Inverted-L

One disadvantage of the 135' horizontal doublet when used on the upper HF bands is that the pattern breaks into a collection of fairly narrow lobes with deep nulls between them. Since the nulls change position from band- to-band, the user is often surprised to discover that signals from certain directions are weaker than expected.

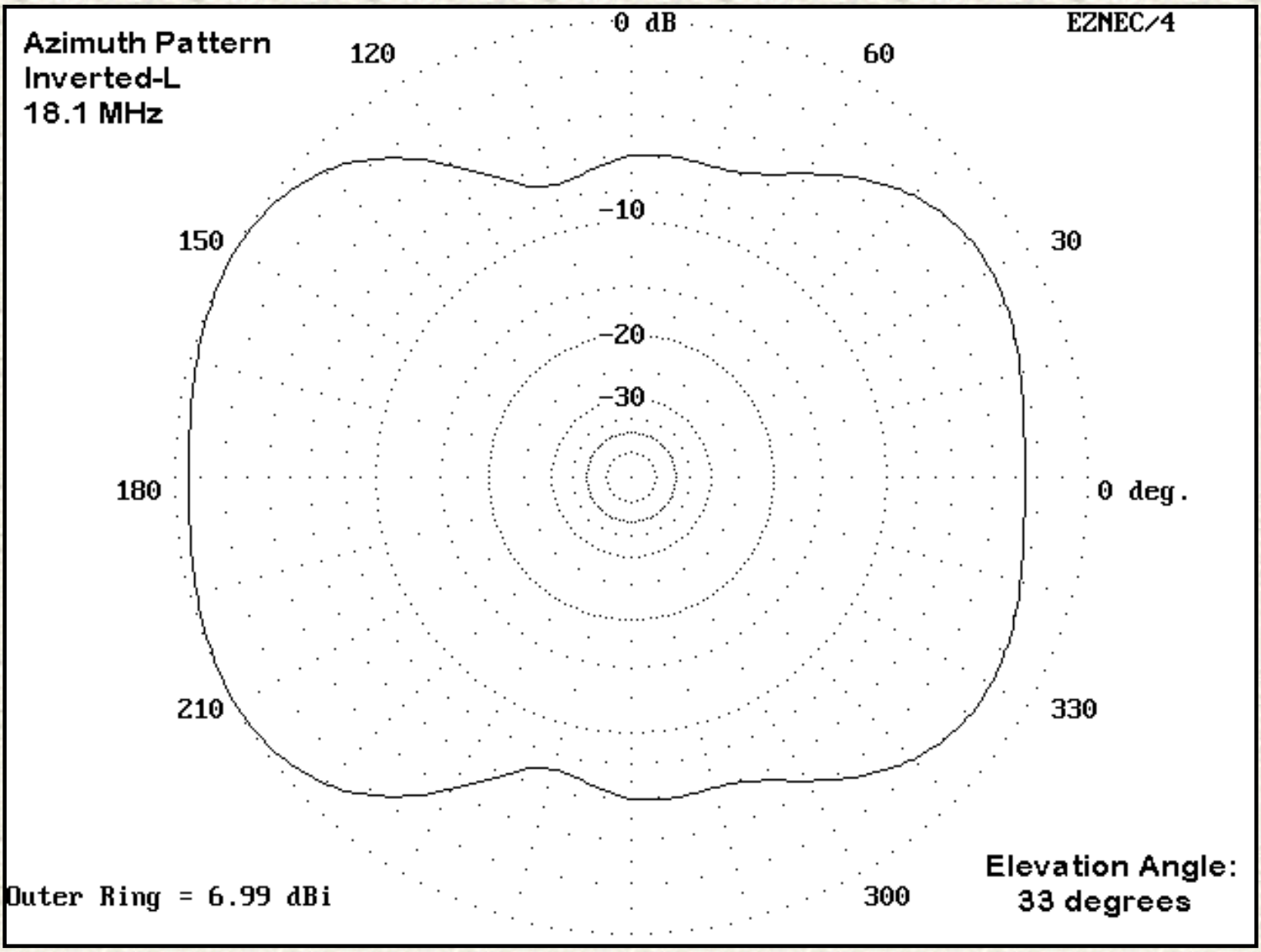
The inverted-L, when fed with parallel transmission line and an antenna tuner, is not wholly exempt from this phenomenon. However, since one arm is fully vertical, the nulls tend to be much shallower. At the same time, gain peaks are less pronounced.

The following table provides a rough guide on what to expect from each of the amateur HF bands:

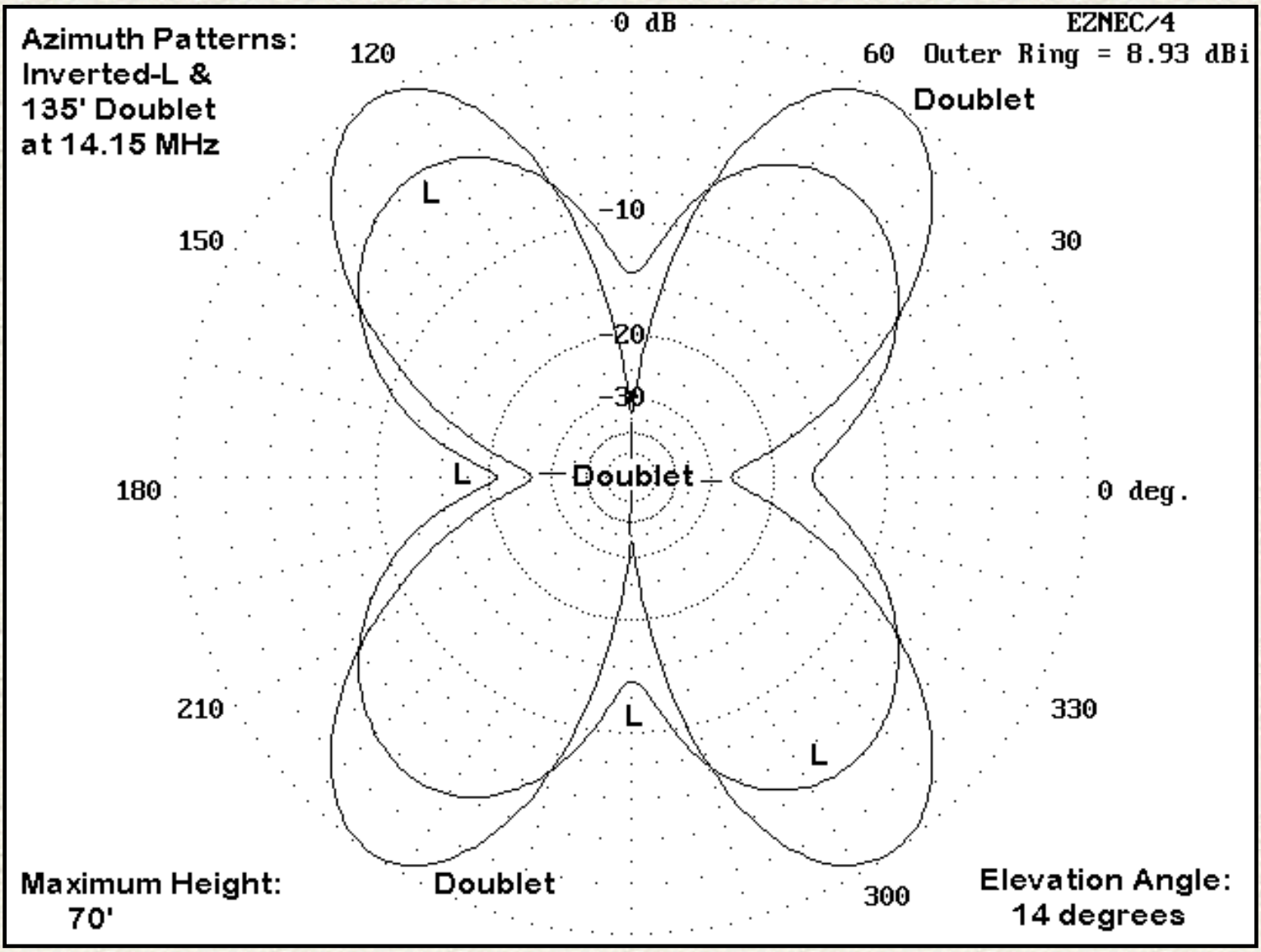
Frequency MHz	Gain dBi	To angle degrees	Feed Impedance R +/- jX Ohms	Pattern Shape (approximate)
3.7	1.93	44	66 + j 4	Broadside oval
7.1	4.09	26	6500 + j 300	Broadside oval
10.1	4.04	20	150 - j 500	Square

14.1	5.38	14	2000 + j2300	4 lobes
18.1	6.99	33	165 - j 255	Square
21.2	6.74	9	750 + j1400	6 lobes
24.9	6.63	8	210 - j 520	6 lobes
28.5	7.55	7	575 + j1000	8 lobes

Even harmonics of the antenna's fundamental frequency show high impedances, in some cases with a high reactive component. The WARC bands show more moderate impedances at the antenna feedpoint. Use of 450-Ohm or 600-Ohm parallel feedline is recommended in order to provide reasonable values of impedance at the antenna tuner terminals. As with all such antennas, if a tuner seems unable to effect a match on a given band, adding a short section of feedline between the existing line and the tuner output terminals will often correct the situation.



The annotation "square" to describe the azimuth pattern is illustrated by the 18.1 MHz pattern. On this band, the strongest signal occurs at the second elevation lobe. There is a usable but less strong lobe at about 16 degrees elevation. Note the absence of sharp nulls and lobes.



Even where lobes and nulls do occur, both are much less pronounced than they are with a standard doublet. The figure shows the differences for the 20-meter band. Doublet nulls exceed -25 dB relative to the lobes, whereas inverted-L nulls are under -10 dB relative to the lobes, which are also broader than those of the doublet. Of course, peak gain of the lobes is about 4 dB less than for the doublet lobes. For some types of operation, but certainly not for all, the absence of strong nulls can be more advantageous than a few extra dB of gain in very specific directions.

The 100' Center-Fed Inverted-L

Many hams who cannot erect a full 135' long inverted-L can often manage a 100' version of the antenna. This length would require 50' of horizontal run and about 55' of height to place the vertical section at least 5' off the ground. As we have seen, higher installations will yield better results, but the present values will provide a kind of worst-case scenario for modeling that antenna. Since the sketches for this shortened version of the inverted-L, which is about 70% full size at 80 meters, would be the same as those for longer versions, we can jump directly to a table of values for multi-band use of the antenna.

Frequency MHz	Gain dBi	To angle degrees	Feed Impedance R +/- jX Ohms	Pattern Shape (approximate)
3.7	1.32	51	30 + j 425	Broadside oval

7.1	3.13	31	305 + j1010	Broadside oval
10.1	4.95	24	2150 - j3100	Broadside oval
14.1	4.64	19	120 - j 185	Square
18.1	5.51	34	965 + j1785	4-Leaf clover
21.2	5.00	12	475 - j1300	4-Leaf clover
24.9	6.07	10	160 + j 95	6 lobes
28.5	7.03	9	1775 + j1990	6 lobes

As one might expect, the shorter antenna breaks into multiple lobes more slowly with increases in frequency. Moreover, the pattern of high and low feedpoint impedances differs greatly from the pattern for the 135' version. Given the lower top height, the elevation angles of maximum radiation are somewhat higher, especially on the lowest bands of operation. (Note that the band on which an unexpected high angle of maximum radiation occurs for both versions also shows a lobe of nearly the same strength at a lower angle--just about 20 degrees lower. Hence, useful radiation occurs on that band--in this case 17 meters.) Shorter antennas--down to about 90' overall wire length can be built and used on 80 meters. Below about 90' overall wire length, the antenna becomes essentially a 40-meter-and-up inverted-L.

Conclusion

The center-fed inverted-L has the potential to be a quite satisfactory all-band wire antenna suited to certain environments. The length can be almost anything about 3/8 wl or longer for the lowest frequency of intended operation. Although the overall gain will be lower for each band than the gain of a horizontal doublet using the same overall wire length, the elevation angle of maximum radiation for the L will be lower than for a doublet with the same top height.

There is little evidence, despite the vertical position of one arm of the antenna, that the inverted-L would benefit from a ground plane beneath the antenna. The actual low-angle gain of the inverted-L will, however, vary with the quality of the soil in the region of reflection at a distance from the vertical arm. All patterns were taken over average soil, and soils that are either poor or better than average will tend to show a higher gain and lower take-off angle, at least on the fundamental frequency.

The electrical lineage of the center-fed inverted-L is from the dipole by way of the inverted-Vee. For the amateur yard that is short on horizontal space but long on tall supports, the inverted-L may be the antenna of choice as an all-band wire--whether used as the primary station antenna or as the back- up for more complex arrangements.

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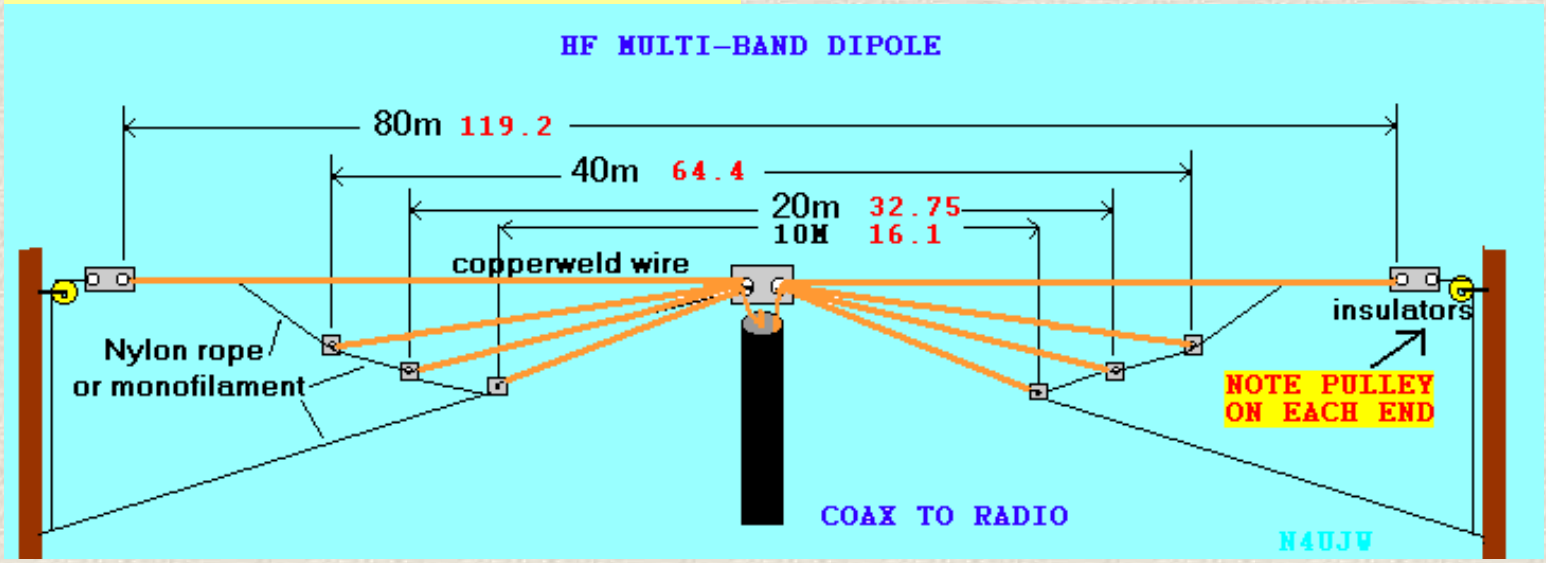
BUILD THIS MULTIBAND FAN DIPOLE FOR ALL BAND HF EXCITEMENT



Here is a simple and easy to build multi band horizontal fan dipole that can be constructed for all band operation from 160 meters up thru 6 meters or even higher.
In the drawing above, it is shown for just four bands, 80 thru 10. One separate dipole for each band needed. However you can build it to suit your own preferences by using the standard formula for a dipole:
468/freq mhz = total length for each band. Use the formula for your desired center frequency.

Each dipole length above in RED is in feet and tenths of a foot for the center of the General portion of each band and is derived from the above formula and should be cut a little longer for swr trimming. USE #12 TO #14 GAUGE COPPERWELD WIRE IF POSSIBLE or use what you have on hand.

Start with your lowest (in frequency) band of operation as the main (top) support for the entire setup. Cut it per the formula and a little longer. Try to use a wire size that will support the other dipoles. This is the main support for all the other dipoles and must carry their weight.
Cut a dipole for each band of operation.
Cut each full length in half....example: for the 10 meter length from the formula you get 16.1 feet for the total length. Cut it in half equaling approx 8 feet per side. Make sure you cut each length about a foot or more longer for swr trimming and attaching to center and end insulators!
If you are building the four band dipole above, you should have 8 lengths of wire scattered all over your work area.
WARNING!
DON'T DO IT IN YOUR LIVING ROOM, THE XYL WILL NOT BE VERY HAPPY WITH YOU AND AFTER SHE GETS FINISHED WITH THE QRM,,,, ALL YOUR ANTENNA BUILDING WILL HAVE TO BE DONE FROM THE DOG'S HAM SHACK! HI!
It is assumed that you have your end support poles, trees, center and end insulators, pulleys all ready to go before you start working on the actual dipoles.

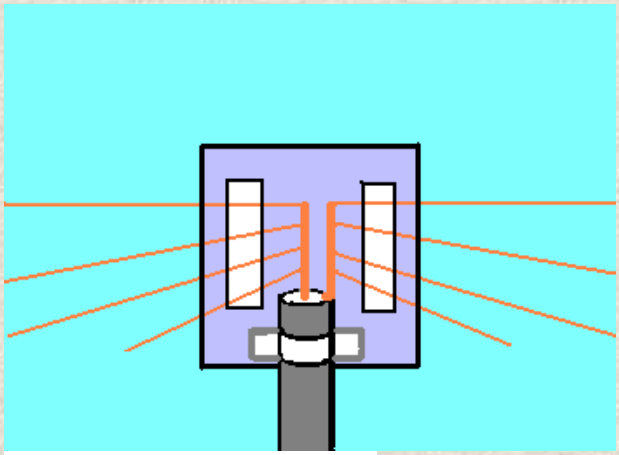
A very important part of this design is the installation of the pulleys (in yellow on drawing) on each end attached to each side support.
They are added to this design due to the swr trimming process and make it very easy to pull the entire antenna up and down while making the swr adjustments. Mount a suitable size pulley on each end attached to your pole, trees, etc for the diameter of cord or rope used to support the system.
Start your antenna trimming with the top dipole.... attach your coax to the center insulator leaving several inches of the center conductor and shield exposed. Each half of each dipole will be connected to the coax center pigtail and the shield separately. In other words,

connect one side of the dipole to the center conductor and the other side to the shield.
Attach the other end of each half of the longest wire to the support cord and run thru the pulley on each end and pull the dipole up into the air between the end supports. Check swr.
Trim as needed with low power for lowest swr possible, lower with pulleys, attach the next highest band dipole electrically to the same point as the first dipole,, raise it to operating height, check swr,,lower for trimming, up and down, up and down.....due the same for all other dipoles for each higher band of operation.
When you are finished with the highest band of operation, pull the entire system up with the pulleys and tie of at the bottom securely.
Make certain that the coax center conductor is attached to one half of each dipole and the shield to the other half. All dipole ends at center insulator are connected together.
This may not be very clear to the new antenna builder so please see the drawing below for the center insulator arrangement.

The white areas in the drawing above are mechanical supports, clamps, wire ties or whatever your genius can come up with to support the main (top wire) and the weight of the coax.
Remember, all the weight of this antenna system is supported by the top wire.
The connections should be soldered and all should be sealed including coax end from water, ice, snow etc.
Use a 1:1 balun like the "Ugly Balun" project page on this site close to the center before coax goes to your rig.

For best performance get it as high as possible and remember that since this is a dipole arrangement, it will be somewhat bi-directional towards and away from you as viewed in the drawing. (BROADSIDE)
Remember that all elements will interact with each other in the tuning process and the final setup must be secured so the angle or distance between each dipole does not change when blowing in the wind, etc
The angle or distance between each dipole is not critical but the final spacing must be maintained!
It will take lots of work (trial and error) in getting each dipole to the lowest SWR. Just keep TRYING.
It should also be noted that the antenna can be used in an inverted v fashion but remember the spacing should be secure in the final operating posistion. Tune it as in all the above instructions.
You may use a tuner with this antenna un-trimmed to save a lot of work!
EXPERIMENT! EXPERIMENT! EXPERIMENT!

BACK TO ANTENNA LAB



<<Little gray boxes are insulators>>

HAM UNIVERSE

MULTIBAND ONE ELEMENT VEE BEAM

BY LA0HV

PETER OF NORWAY

A Six Band One Element Beam!

20 meters thru 6...cool!

Hi N4UJW.
Thanks for a GREAT web site !!!!!!!!
"I sent this email to Darrell, have a look, I think you will like it!"

Vy 73 LA0HV, Peter

YES PETER, WE DID LIKE IT!.....HERE IT IS FOR ALL HAMS TO BUILD...ENJOY!!

THE ANTENNA!

"First of all THANKS for the great web site of the "ONE ELEMENT V-BEAM FOR 15 METERS from Darrell, KB4XJ!

It really turned my antenna farm upside down! And what a relief.
My wife loves me again!
It reduced the antenna farm from several Quads and 3 element junior beams for several bands into the latest one element design.
A 6 band one element horizontal V-beam covering 20, 17, 15, 12, 10 and 6 meters!
Giving from 2.1 dBd (thats over a dipole) on 20 meters to 10.8 dbd on 6 meters."

Design is simple:

Take a Double Extended Zepp design for 6 meters using 1/2 inch or larger element material such as EMT, copper or aluminum. Copper is heavy! DO NOT USE WIRE (design freq. 51.110 MHz - "downunder DX window")
Each element will be 12.33 ft per side.. Feeder must be 400-600 Ohm ladderwire for at least 2.64 ft. (See Design calculator below)
Mount it in a 90 degree angle just like the "KB4XJ" Horizontal V-beam shape.
(I have not experimented with any other angles)
Now connect a 4 ft 400 Ohm ladderwire.

Connect a 4:1 CURRENT balun at the end of the ladderline (balun design must cover 6 meters)
Connect a remote tuning unit directly to the balun (ATU must cover 6 meters)
for example, the LDG RT11 ATU. A GREAT TUNER!
(Click here for info. The LDG ELECTRONICS RT11 ATU). Click back here when done.
The ATU is not a good solution for "normal" beams, because you cant tune all the other elements so easy.
But this is a ONE element beam so it WORKS GREAT!
Connecting a tuner directly at the feedpoint (the balun) will give a TRUE MATCH, not just a happy transmitter.
(Dont use a "by the rig" tuner, it won't do well).

Look at the dimensions and GAIN of this antenna:

On 20m: One element + wire is 1/4 wave. Gain: 2.1 dBd. Fb > 18 dB
On 17m: Almost a 1/2 wave dipole. Gain: 2.8
On 15m: 1/2 wave dipole. Gain: 3.3 dBd
On 12m: A little longer: 4.0 dBd
On 10m: The antenna is a little out of band but the Balun and RT11 tuner does the job. Gain: 5.8 dBd
On 6m: EDZ v-beam. Gain: 10.8 dBd. Fb > 25 dB

ALL WITH ONE ELEMENT !!!!!!

Now isn't that cool!

Vy 73 LA0HV, Peter
Horizontal V-Beam Diagram

LA0HV EXPANDS ON A GREAT DESIGN!
THE BELOW INSTRUCTIONS WERE TAKEN FROM VARIOUS EMAILS FROM PETER, LA0HV with added comments by the original designer of the
ONE ELEMENT VEE BEAM
Darrell, KB4XJ
Some contents edited for clarity.....excuse any spelling errors...Engish is not his native language!

Here is the basic theory:

**The triDouble dipole has two elements of 3/4 wavelength each.
It has resonance at the design freq but the impedence is high (500 Ohm) and the reactance/inductance value changes dramaticly when you change frequency.**

**To compensate for this you shorten the element length a bit and add openwire, 400-600 Ohm (not critical). This will decrease impedance to 200 Ohm at the design frequency.
This is why I use a 4:1 Current balun (must be a current balun).
This is the Extended Double Zepp, EDZ design parameters:**

**Length of each element: 0.64 x l
Minimum length of ladderwire: 0.137 x l**

**Now the great side effect is, that the antenna will work as good as, or better than a dipole at any frequency higher than the frequency that match a 1/4 wavelength at one of the elements.
Now using an open wire feeder, it will also work at the lowest freq and up, corresponding to a 1/4 wavelength, when you add the element length to the wire length. But now of course, gain will decrease because the ladder wire will be a part of the antenna.
Now the EDZ will be easily matched with a remote ATU that can handle rapidly changing reactance/inductance values. A remote ATU that can handle swr of 1:10 will do the work easily.**

Back to the V-beam design:

**The EDZ-V-beam with design freq at 51.110 MHZ will work with very high forward gain at the design freq, and as a normal "KB4XJ" V-beam at the 15 meter band (look at the element length).
Using open wire feed, minimum 4 ft
(longer is ok, but remember it is a part of the antenna at frequencies under 21 MHz), will give an easy match from 14 MHz and up.**

Conclusion: The trick is to use openwire, a 4:1 balun and a Remote ATU like the LDG RT11 TUNER connected directly to the balun.

**Vy 73, Peter
LA0HV**

THE LATEST INFORMATION

EMAILs RECEIVED AS OF (11-12-02) from Peter LA0HV and Darrell KB4XJ

Hi Darrell & Don.

I just finished the last experiment of the year. It's snowing heavily now in Norway, 1 ft per hour! Minus 10 deg Celcius .... Brrrr

Last experiment revealed something very interesting about the current in the elements of the V-Beam:

I made the KB4XJ 15m V-Beam design using thin wire, same length and angle.
This happend: Field strength measurement now showed a major backloop and a decrease of forward gain (almost no front to back). You can say that the Beam effect almost disappered.
I then took 1.25" kopper (copper) tubes, same length and angle, and all of a sudden I got a V-beam again. Backloop gone, forward gain increased to almost 6 dB !!!!!!!!!!

So .. conclusions from a freezing cold, windy Norwegian Hill-top is:

A "fat" element is turning the design from a resonant V-wire antenna into a Aperiodic (stable current) element V-Beam. The thick elements are doing the same job as a terminator in a Rhombic design . WOOOW !!!!!

Freezing smiles from LA0HV.

Vy 73 Peter

From: Darrell & Kay Koranda [mailto:kb4xj at strato.net]
Sent: **9. november 2002** 01:16
To: Peter Grün
Re: Developing a 6 band Single element horisontal V Beam

Hi Peter,
The Field Strength Meter and getting readings.
I checked across the back side of the V beam and showed a field strength reading of a .1. At the sides of the front ends power rose .5 and just inside the ends power rose to a 1. In the near center power rose 2. and in the dead center it was a 3. In just the field strength reading I was figuring around 28 db difference between front to back. In the tests I made I was using 10 watts of power. The antenna performed very good 2 years ago when I first truly tested it on field day 2001. Field day 2002 was not as good, problems with generator, blowing rain. Anyway back to the Horizontal V, in my early study on long wire antennas it was noted that power was related to one wire and that when a wire had a twin, the power doubled. Further study of log periodic dipoles (when a log dipole is bent forward it will have an added gain of 2.5 db., it was also found the beginning horizontal V was a 1/4 wave. Now in my

experience I found that induction between the elements changes the resonant frequency and lowers the frequency of a dipole and as a result you have to add 2 inches per side to make up for the induction factor. In your 90 degrees configuration, the higher you go in frequency the broader the front lobe should be, in that a horizontal V at 90 degrees at 1/4 wave on design, at longer wavelengths the V angle can be narrowed down, but the antenna does work and I wonder why it was never truly correctly reported, not having the works of Dr John Klause at hand. I spent 4 months researching the horizontal V antenna, and I spent over a month refining the center mount and have the antenna and mount down to a lean 7 pounds! Darrell KB4XJ

==========

THESE LAST TWO EMAILS ARE UN-EDITED!

----- Original Message -----
From: Peter Grün
To: 'Darrell & Kay Koranda'
Sent: Friday, November 08, 2002 9:51 AM
Subject: SV: Developing a 6 band Single element horisontal V Beam

Hi Darrell.

The pointing right up idea of yours is also greate. I once did that experiment with a 2 element for 80 meter. For a long time I irritated me that the G5RV did'nt preform wel. I found out that when you are using a Inverted V type antenna over a sandy or rocky soil, it dosent work (I live on a rocky hill). The ground is absorbing the E-field. The reason why the Inverted v type antennas like the G5RV works for most people, I think, is because the Earth with a good ground-mirror soil works as a FireBack reflector.
So I took a 80meter dipole and mounted a reflector between the dipole and ground, the "beam" pointing directly in thy sky. The effect was thremendous! Within a 1000 miles range nobody understood how my signal could "hit" them with such a ekstreme powerfull strength. It was a 80 meter contest winner antenna. I think that it also is a good DX antenna, the only problem is that stations within 1000 miles is "bending" your S-meter, so you really need ekstremly strong filters to be able to work in the DX-window.
Back to your V-beam: I dont understand how it is possible to actually get a good FB-ratio on the V-beam. I did some fieldstrength readings and I am still suprised!!! It's so cool! Now talking about "cool" winther is closing in on me here in Norway so I have to wait until april for more eksperiments.
Havent got any RTTY at the moment, but wil get some equipment during the winter, I wil look for you on 15 then.

Vy 73 de LA0HV, Peter

-----Opprinnelig melding-----
Fra: Darrell & Kay Koranda [mailto:kb4xj at strato.net]
Sendt: 7. november 2002 13:02

Til: Peter Grün
Emne: Re: Developing a 6 band Single element horisontal V Beam

Hi Peter, Did you do any Field Strength readings yet? That is what shocked me with the orginal design of the horizontal V design on 15 meters. The very high increase in forward field strength in the antenna and the decrease to the rear, could hear 360 degrees but could see a noticeable increase on the S-Meter when you rotated to the right direction. It really nice that the antenna can be rotated with a cheap TV rotator, I'll have to try your modified version. I was thinking of trying another experiment but haven't had the time to do it yet, I was thinking of pointing the V straight up and using the sky to bounce off of or launching at a 30 or 40 degree and up, of the two experiments I think the second may be of a better. In the First your pulling the signal straight down out of the sky and shooting them back up and scattering, in the second your pulling signals down in a direction and shooting back in the same direction. The First pointed up has the advantage of making your antenna appear to be say 75 miles up with nearly omni coverage. The Second pointed off with a 30 or 40 degree rise should provide longer distance. This is what I think will happen, the extended double zepp is basically a 3/4 wave antenna shorted to 5/8 wave with a 1/8 wave of ladder line, I have one but haven't tried it in the V configuration yet. I have time constraints and have to go off and work, I started my own business in refrigeration repair and also work with a solar pool heating contractor and I have to get going to work, we have a 40 panel system to put back on a building and we spent 6 hours yesterday just putting back up 4 pieces of pipe, and it looks like another 2 days on the job. If you operate rtty, I operate around 21.278 Mhz, current project is connecting my new radio a Patcom PC 9000 to my Pakrat 232 TNC and then to a old laptop for some portable rtty with the One Element Beam with Solar Power. Have yet to get the charge controller for the solar panel, I'am using a string of silicon diodes to drop the voltage to 13.8 volts DC. Got to Go good by for now, hope we can exchange information later. Darrell KB4XJ 73's

BACK TO ANTENNA LAB
ONE ELEMENT VEE BEAM PROJECT

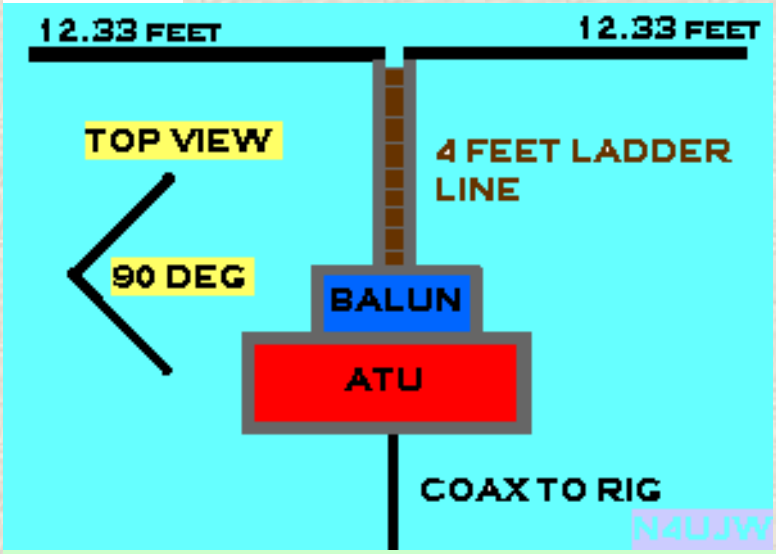
MHz

= ft. (or inches.)

USE THIS CALCULATOR FOR EXPERIMENTATION!
>>>>
or use these formulas
one leg = 630/freq mhz = ft
ladder line = 135/freq mhz = ft

EMAIL
DARRELL
KB4XJ

EMAIL
PETER
LA0HV



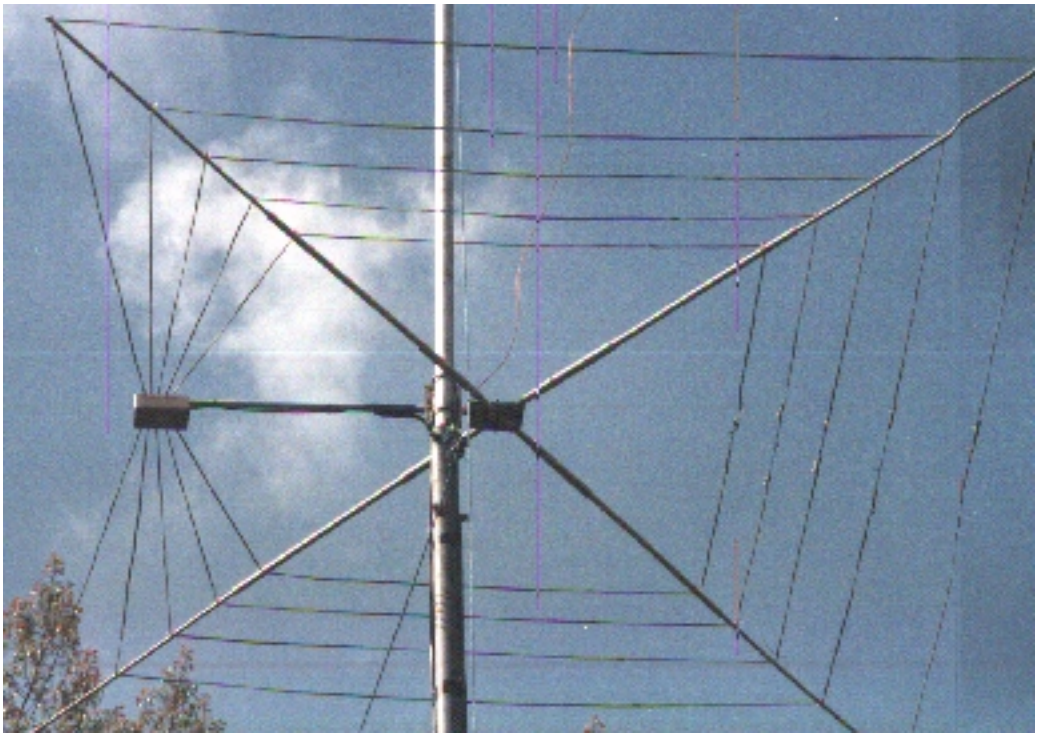
EDITORS NOTE from N4UJW:
"In the true spirit of Amateur Radio, Peter saw the original plans for the ONE ELEMENT BEAM on this site and saw that an improvement could be made to this fine design which you have just seen AND HAS SHARED IT WITH ALL OF US!

Can YOU improve on his improvements or have a good antenna design or antenna project that is hard to find on the internet? Please send them to us!" email here

EXPERIMENT! EXPERIMENT!..... 73 N4UJW

Peter Grun
LA0HV
Svinndalveien 96
Skiptvet, 1816
NORWAY
License Class: A

The Cobwebb HF Antenna



The G3TPW Cobwebb antenna covers five bands, 14 - 28 mhz, including the WARC bands. It is made by SRW Commuications Ltd (Steve Webb), Astrid House, Swinton, Malton, N. Yorkshire YO17 0SY (tel: 01653 697513).

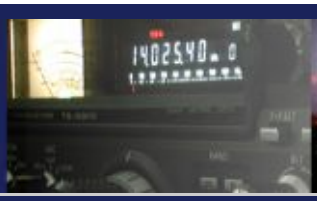
It is strongly made using fibre glass rods and comes pretuned, but is easily adjusted to one's own frequency of preferrence if required.

It consists of five nested dipole balun fed and mounted horizontally. Rotation is not required. Fed with 50 ohm coax. Can be used indoors in a loft as it only measures eight feet across and is very light to handle.

I use one here ar G3YCC mounted on a pole at the bottom of my garden, see photograph. Note, in my case the pole passes through the centre of the Cobwebb, but it works very well and I have had a QSO on SSB - rare for me! - with VK on 15m through contest QRM.

Although I have no commercial connection with Steve, a professional aerial designer, I have heard him speak on antennas at our local club and can recommend the Cobwebb, particularly for those with restricted space. I have a friend who uses one in his loft, which enables him to get onto 5 bands, which otherwise he could not do, owing to available space outside.

Back to the first page



YANKEE CLIPPER CONTEST CLUB

Double-L Antenna For 80/160

Don Toman, K2KQ
toman@ibm.net

A popular misconception about vertical antennas for the low bands is that they must have elaborate ground systems. Here's a vertical antenna for 80 and 160, fed with a single feed line that is simple, effective, and requires no ground system. You won't beat the 4-squares, but you will hold your own against a grounded quarter wave with ridiculous amounts of copper in the ground.

Rather than get into the theory of why this antenna works, I will simply describe it here and let results speak for themselves. If there's a demand, I'll do a follow-up article on the relevant theory.

The antenna is a center-fed half-wave vertical with about 70 feet of vertical length with the remainder of the top and bottom of the antenna bent horizontal and parallel to each other. The antenna looks like a squared-off letter "C" fed in the middle of the vertical part.

Thus, the 160-meter antenna is a 270-foot dipole fed in the center with the bottom antenna wire bent parallel to the ground about 10 feet off the ground and the top at 80 feet off the ground. The horizontal parts are 100 feet long and parallel to each other.

The 80-meter antenna is a 130-foot dipole, fed in the center with 70 feet vertical and 30 feet horizontal 10 feet off the ground and 80 feet off the ground.

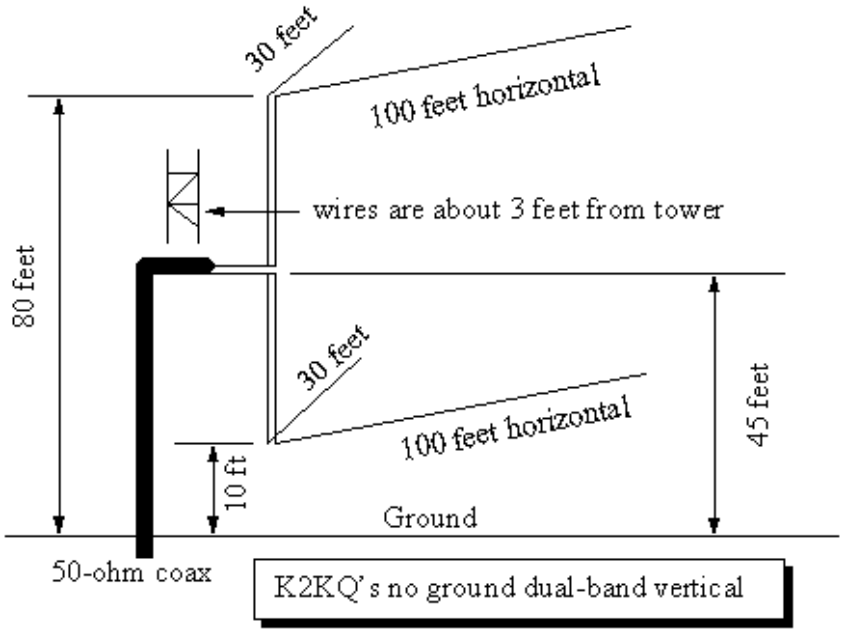
Think of it as an inverted L fed against an L. The two Ls are balanced with respect to each other and because the currents in the horizontal sections are out of phase, the antenna has a minimum of horizontal radiation.

The 80 meter and 160 meter antennas are separate, fed from a common coaxial feed line.

In my case, the 80-meter and 160-meter horizontal sections are about 30 degrees apart. The 160 horizontal wires run east while the 80 wires run east-northeast.

On both bands, the bulk of the current flows symmetrically in the center of the antenna, with the current peak about 45 feet off the ground at the feed point. On 80, the current loop peak is about 0.16 wavelengths above ground and on 160, the current loop is about 0.08 wavelength above ground.

The accompanying figure illustrates the antenna.



You can adjust the resonance of the antenna by adjusting the lengths of lower horizontal sections. The small asymmetry doesn't bother anything. The center impedance of mine at resonance is very close to 50 ohms on 160 and close to 70 ohms on 80. The 160 antenna presents high impedance at 80 meters and the 80-meter antenna looks like a parallel capacitor across the 50-ohm 160 antenna. The 160 antenna covers 1800-1860 with under 2:1 SWR. I needed to take about 2 feet off the horizontal section to get mine resonant at 1830. If 80-meter current flows in the 160 antenna, it tends to flatten the current loop in the vertical section.

The 80 meter antenna is resonant at 3750 with the 130-foot length shown. The VSWR is under 2:1 over the DX part of the 'phone band. It needs to have some length added to cover the CW portion. I haven't tried to bring it to resonance in the CW band, but have chosen to feed it through a tuner.

I originally had this antenna hung from trees. This year I put up an 80-foot Rohn-25G with three sets of guys. I hung the wires from ropes attached to the tower so they are separated from it by about 3 feet. The coaxial feed line comes off perpendicular to the antenna and is then taped to the tower. The center conductor goes to the top and the shield goes to the lower part. Before the coax turns on to the tower, I've wrapped some 30 feet of it into a coil. I expected to see a lot of interaction, but the tower and guys seem well off resonance at the operating frequencies and I didn't see any to worry about.

The first QSO on 80 was VK6LK, long path on SSB at sunset on September 12. I've worked a few ZSs, HF0POL, LU, and the usual Europeans. On 160, the first QSO was KP4SN on September 15. In the couple of weeks since hanging it from the tower, I've worked ZS6UT, TU2MA, TL5A, VK6VZ, VK6LK, VK3ZL and NL7Z and the usual horde of Europeans on 160 with no fuss. I run about 800 watts out from a Ten-Tec Centurion.

Last January, I had about 200,000 points with 750 QSOs in the CQ WW CW 160 contest using the predecessor hung from trees.

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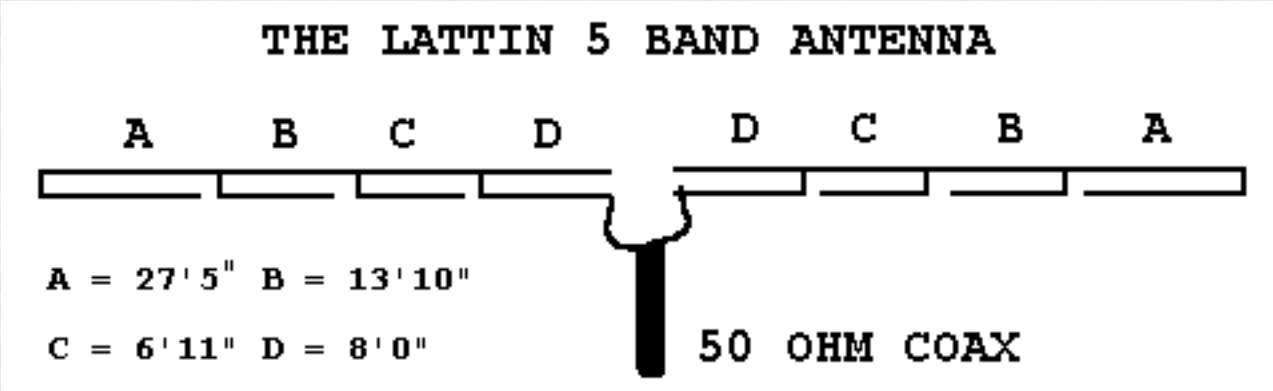
THE LATTIN 5 BAND ANTENNA

(DESIGNED BY W4JRW)



THE LATTIN 5 BAND ANTENNA (FOR 80, 40, 20, 15 AND 10 METRES)

The antenna was named for **W4JRW** who invented it and holds a patent on the basic principle and uses quarter wave stubs, which act as insulators at the frequency for which they are cut. For example, the 6'11" stub (quarter wave times the velocity factor 0.8 of the feed line used) blocks RF for 28 mhz from reaching the rest of the antenna.

In the example shown in the diagram, tubular foam filled 300 ohm feed line was used, which has a VF of 0.8. Other feedlines may be used, for example, slotted ribbon and the length of the stubs worked out using the correct velocity factor

BUILDING THE LATTIN ANTENNA

This will require some forethought and planning. Avoid cutting the continuous top wire, which supports the whole system. I wonder if it might be an idea to use a suitable polypropylene line to support the wire, which may be subject to breaks, especially at the solder points?

A suitable centre piece may be constructed and constructors may want to include a balun at the centre of this balanced antenna, which is fed with unbalanced line (coax).

If you do have a go at this, please let me know - I have always wanted to make one and never as yet got round to it!

A version of the Lattin could be designed for all bands, including the WARC bands - get snipping!

Frank, G3YCC.

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Trapped Antennas

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Additional information on this subject and related topics can be found in back issues of *QST* and the following:

The ARRL Antenna Book

The ARRL has an extensive catalog of books and materials related to Amateur Radio.

Articles

Note: Some of the following articles are in Adobe Portable Document Format (PDF) files. To view and print these files, you'll need a copy of Adobe's Acrobat Reader program. (Version 3.0 or later required). More information here.

- Coaxial cable Antenna Traps (1,318,022 bytes, PDF file) MEMBERS ONLY
  QST May 1981, pp. 15-17
  Both the coil and capacitor of a parallel-resonant antenna trap can be made from the same length of coaxial cable. Sound intriguing? Here's how.
- Build a Space-Efficient dipole Antenna for 40, 80, and 160 Meters (570,975 bytes, PDF file)
  QST July 1992, pp. 35-36
  A new trap design, using only RG-58 and PVC pipe, yields better space efficiency than conventional coaxial traps.
- Two New Multiband Trap Dipoles (785,934 bytes, PDF file) MEMBERS ONLY
  QST August 1994, pp. 26-29
  W8NX details a new coax trap design used in two multiband antennas; one covering 80, 40, 20, 15, and 10 meters, and the other covering 80, 40, 17, and 12 meters.

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- An Improved Multiband Trap Dipole Antenna (156,121 bytes, PDF file) MEMBERS ONLY
  QST July 1996, pp. 32-34
  You need this - traps with lower loss, higher Q, increased power-handling capability and four-band coverage. Also build a multiband dipole for 80-, 40-, 17-, and 10-meters only 84 feet long.
- Taming the Trap Dipole (151,344 bytes, PDF file) MEMBERS ONLY
  QST March 2002, pp. 28-30
  A trapped dipole for 10/15/17 meters

guide to four VoIP systems: EchoLink, IRLP, eQSO and WIRES-II.

APRS -- Moving Hams on Radio and the Internet -- **Now Shipping!** -- A Guide to the Automatic Position Reporting System

Note:

Contact information for suppliers mentioned in the above articles should first be confirmed using **TIS Address Database Search**.

Bibliography (Members Only)

ARRL Periodicals Index Search - This database contains the *QST* index from 1915 to the present and the *QEX* index from 1981 to the present. For *QST* issues from 1970 to the present, and some selected articles back to 1922 (when construction articles featuring tubes began in earnest), identifying keywords have been added to the technical articles. By entering keywords (ANTENNA) or combinations of keywords (CONSTRUCTION ANTENNA VERTICAL HF) into the **Title words:** field, you may create dynamic bibliographies.
Technical article KEYWORD list. Hints for more successful searching

Suggested keywords for more articles like the ones on this page are: CONSTRUCTION TRAP ANTENNA

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+++ DF4SA amateur radio website +++

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Spiderbeam has now earned its own site.

It has been moved to: www.spiderbeam.net

A TRI-BAND 2 ELEMENT PORTABLE YAGI

After operating QRP during the 1996 ARRL Sweepstakes Contest I decided that the next time I entered a QRP contest it was going to be with an antenna that had gain over the dipole antennas I was using! Operating from a temporary location, a tower and tri-band yagi constructed of aluminum tubing was not an option. After much thought and a lot of reading, I decided to build a tri-band 2 element yagi using wire elenments. The results of this experiment are contained in an article published in the November 2001 issue of QST. A PDF file of the article as well as the EZNEC antenna files is available at the following links.

Portable 3-Band Yagi Article
EZNEC zip file at QST

ADDITIONAL NOTES:

Assembly

When you are ready to assemble the array, attach the 20 meter reflector and the driven elements first and then hang the array between two supports, (trees etc.) at about 5 feet above the ground. Pull it tight so that the array is fairly flat. It won't stay flat because the driven elements are heavier than the 20 meter reflector so support one end of the 2 x 2 end supports on a rung of a step ladder or box so that the array lies horizontal. Add the 10 and 15 meter reflectors next using a little less tension than the 20 reflector. Next attach the feed line. The last step is to adjust the V slings as mentioned below so that the antenna is balanced in the horizontal plane. All adjustment for lowest SWR, if needed, must be done with the antenna raised to its typical operating position. Do not attempt to adjust any of the element lengths or the hair-pin match for lower SWR while the array is close to the ground. Check the SWR on each band with the antenna at a height of at least 25 feet before attempting any adjustments.

V Slings

There are two V slings, one on each end. The secret is that they are not equilateral in shape. The combined weight of the driven elements and feed line is heavier than the reflectors. If the length of the sides of the V sling are equal, the array will turn with the 2 x 2 end supports positioned vertically. Increasing the tension on the side of the V that is attached to the end where the driven elements are attached causes the whole array to turn toward the horizontal plane. Do this by shortening the length of the side of the V that is attached to the driven elements. It is quite easy to adjust in the field. Once the V sling is adjusted, the array stays balanced. You can change direction 180 degrees by pulling on the the feed-line. If you pull it hard enough, the whole array will begin to turn-flip over. Stop it from turning too far by holding on to the feed line once the array has swung over to face the opposite direction.

Balun

It is not clear what to do with the coil of coax at the feed point in Fig. 1. This is meant to be a choke balun to choke off any RF from flowing along the outside of the coax shield. It is best to let the coax feed line drop straight down from the centre insulator, attach it to the centre of the hair-pin shorting bar with a plastic tie or string and then make the balun by coiling up the coax 6 to 8 turns with a diameter of 4 inches or so just below this point. The centre of the shorting bar is neutral potential so there is no problem mechanically attaching the feed line to it for support.

LATEST UPDATES:
November 15/03, RELAY PROBLEMS; see Homebrew Page.
March 15/04, 100 Watt HF Power Amplifier Completed, info to follow.
April 5/04, New addition to link page.
July/August 04 QEX, see article on how to fix your HP8640 if it has no power output.
August 23/04, TOWER COLLAPSED, being fixed; see Antenna Page.

Welcome to VE7CA's Homepage

- **Operator:** Markus Hansen.
- **Location:** North Vancouver, B.C. Canada - Grid CN89.
- **Activities:** Contests, LF-HF-VHF-UHF, Homebrew, Antennas, CW, QRP

Biography:

I obtained my ham radio licence in 1959 at the age of 16 and was assigned the call VE7BGE. (My first Ham Station) After high school I obtained my 2nd Class Commercial Radio Licence at Room 19 in the City of Vancouver, BC. Upon graduating from Room 19, I obtained work as a commercial radio operator and weather observer at Alert BC off the west coast of Vancouver Island at a Loran Station. I was later transferred to Smithers BC, then served for one year at Alert NWT. After returning from the arctic, I was transfered to Terrace BC and finally to a Canadian weather ship named, The Stonetown, stationed off the West Coast of British Columbia. In 1967 I decided to return to school and enrolled in general sciences at the the University of British Columbia, (UBC). While at UBC I was a active member of the university ham club called HAMSOC. (Hamsoc FD 1967), (1986 HAMSOC Reunion), (1992 HAMSOC Reunion), (1995 HAMSOC Reunion). During my second year at university I met my future wife, Gaye. We were married in 1969. We have one daughter and two sons. Our daughter and youngest son are married and we are blessed with two grandchildren. In 1979 I was issued a two letter call, VE7CA.

I have been working in the commercial real estate industry for the last 30 years and have no plans to retire. Here I am in front of my operating desk looking for you on my homebrew HBR2000!

I continue to be active operating mostly CW, chasing DX, looking for new grids on 6 meters and entering contests under the QRP category. I have a strong interest in antennas and have published 4 articles on antenna designs that I have developed. Homebrewing is something I really enjoy as well and lately I have been absorbed in building a high performance HF transceiver. See the above menu for further details regarding my antenna articles and homebrew projects.

Antennas:
160/80 trap dipole, switchable to Vertical T antenna.
Triband Dipole at 35 feet.
Portable 2 Element Wire Tri-band Yagi, See QST November 2001
Portable 2 Element 6 meter Quad
Portable 3 Element 6 meter Yagi
Cross Polarized 144 and 430 Mhz Hi-Gain yagis.

Ham Gear and Test Equipment:
Homebrew 160-10 Meter High Performance Transceiver HBR2000
Homebrew 6 meter CW/DSB 10 Watt DC Transceiver
Modified Norcal Sierra QRP CW Transceiver 40-10 meters
Homebrew 40 Meter Optimized QRP CW Transceiver based on W7EL design in August 1980 QST
Homebrew 160-10 Antenna Tuner

Kenwood TS-940s HF Tranceiver (FOR SALE)
Yaesu FT-726R VHF/VHF Transceiver
Homebrew 6 Meter 70 Watt Solid State Amplifier
Heathkit SB220 Linear
Homebrew Spectrum Analyzer based on W7ZOI/K7TAU design, August 1998 QST
Homebrew Personal Network Analyzer based on design by Steve Hageman, see Jan/Feb 1998 QST
Homebrew Power Meter by W7ZOI/W7PUA, June 2001 QST and K3NHI additional features May/June 2002 QEX
HP8640B signal generator. See July/Aug 2004 QEX if your gen. has no output power.

You are visitor 04263 since Aug 26, 2002

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Another web page by;

BUX COM ®TM

"Home of the RASCAL"

My Favorite Multi-Band Antenna

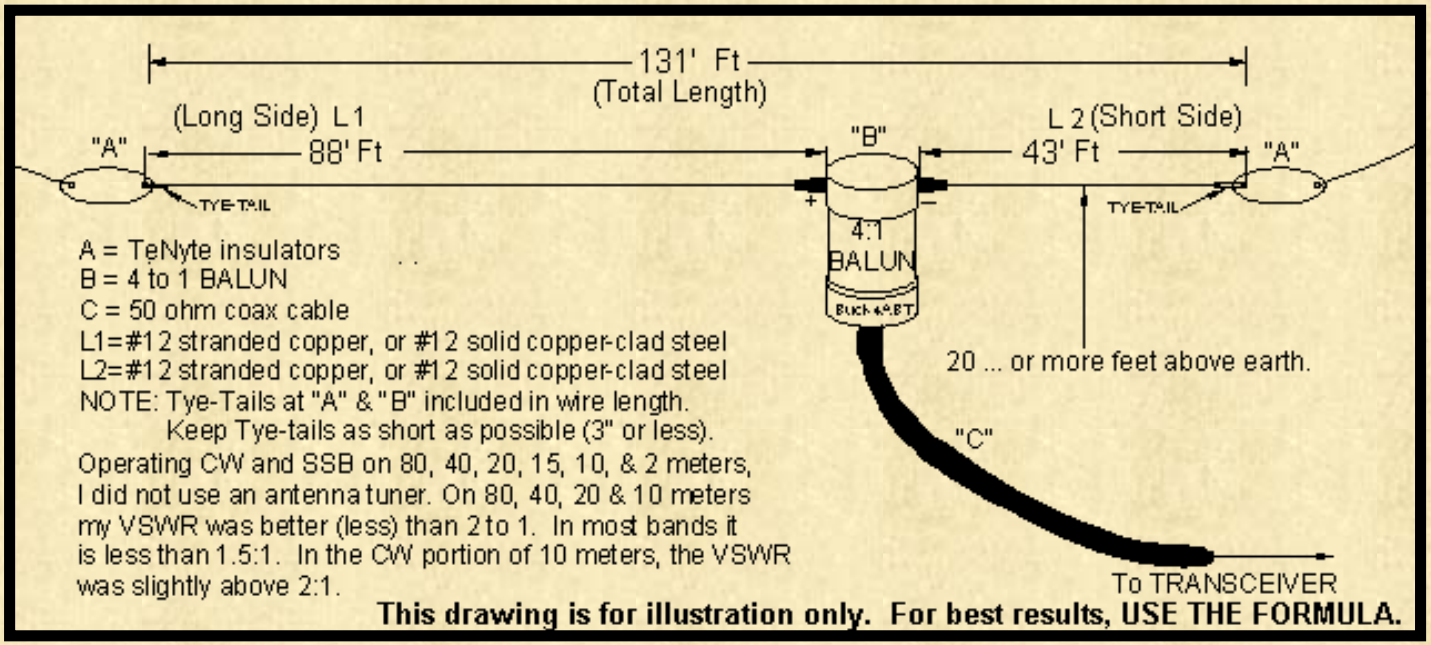
The "WINDOM"

By G. E. "Buck" Rogers Sr; K4ABT

In September of 1949, I was tired of climbing poles and trees to move, remove, add, or change my single-band HF antenna's. In those younger years of my HAM radio hobby, I had used single band dipoles and doublets for almost every HF Amateur band. I had tried long-wires, doublets, dipoles, and Zepps, but again, operation was restricted to single band operation, maybe two bands at most.

I had heard of the "Windom" and read a few articles about the Windom, but most of my thoughts were ... ho-hum.. just another dipole fed a bit off-center. Then one evening at a meeting of the GARC in the old "Sea Scouts" club house near the Coosa River in Gadsden, Alabama; I heard some of my "elmers" Gilbert Watson (SK) W4PAC, Gale Caudle (then W4CFB), Jack Kennamer (SK) W4YPC, Bob Bynum W4USM, Austin (Vic) Vickery, Walter Damkohler (SK) W4EBO, W4CWF, Ed Elkins (SK) W4CDI, Homer Dupree (SK) W4OZK, Jim Runyan, Homer O'Dell, Robert Martin, ... and others discussing the Windom *all-band HF antenna*. It was when Jack (W4YPC), mentioned ***using one (Windom) antenna, on the all HF bands*** *that my ears went directional!*

That last phrase caught my undivided attention. "most all HF bands, ...etc"
What ! *A multi-band HF antenna ?* Surely I had been blessed.



To think that I could put up a Windom, and no longer have to climb the poles and trees to hang another (single band) HF antenna was great news to me. To be able to use it without an antenna tuner was icing-on-the-cake. For a kid without extra funds, an antenna tuner was a luxury that I could *not* afford. Even my transmitter was a single 807 rig I homebrewed on an old *Atwater-Kent* radio chassis, my grand-father had given me.

In those days, a BALUN was unheard of. My Elmer's described, a means of connecting the coax to the off-center fed antenna using a nine (9) turn coil of the coax feed-line at the feed point. This coil of feedline coax formed a "de-coupling" loop. The de-coupling loop provided a crude means of matching the feed coax to the antenna, and at the same time, it would reduce the "re-radiation" (RF currents) along the outside (shield) of the feeder coax.

Today we have toroid cores and BALUN devices that provide a more efficient means of coupling RF energy to the antenna (reducing the VSWR, "standing-waves"), while performing better impedance matching. In the drawing shown above, I've drawn the exact dimensions of the Windom I built in 1949. The only differences in my Windom of 1949 and today are:

1) the material the insulators are made of, and
2) I've substituted a 4 to 1 BALUN for the 9 turn, 8 inch diameter, decoupling loop.
AN UPDATE:
Since writing this article a few decades ago for a major HAM radio magazine, I've received tons of mail (and eMail) asking for more information, especially with regards to a 160 meter version; Here then is "the rest of the story."

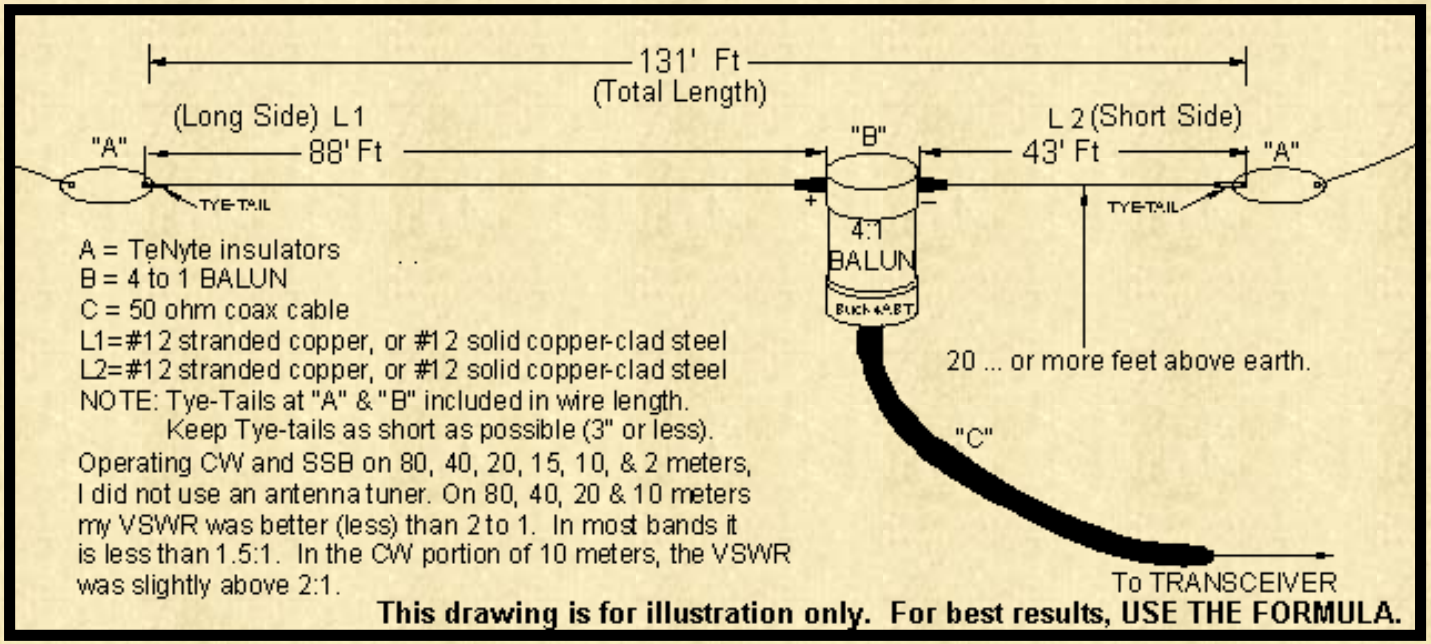
First of all, we'll address the formula, and how to determine the length(s) of each section, using the same old formula that I used in 1949.

Long side.... = 468, divided by the frequency, then multiply by **.**64 (= Feet)

Short side.... = 468, divided by the frequency, then multiply by **.**36 (= Feet)

THUS; for 160 thru 10 Mtrs.....
Long Section; 468/1.8 MHz = 260 x .64 = 166.4 feet.
Short Section; 468/1.8 MHz = 260 x .36 = 93.6 feet.

For 75 thru 10 meters do similar math to arrive at/near the dimensions shown in the drawing below, but for best results, use the formula.

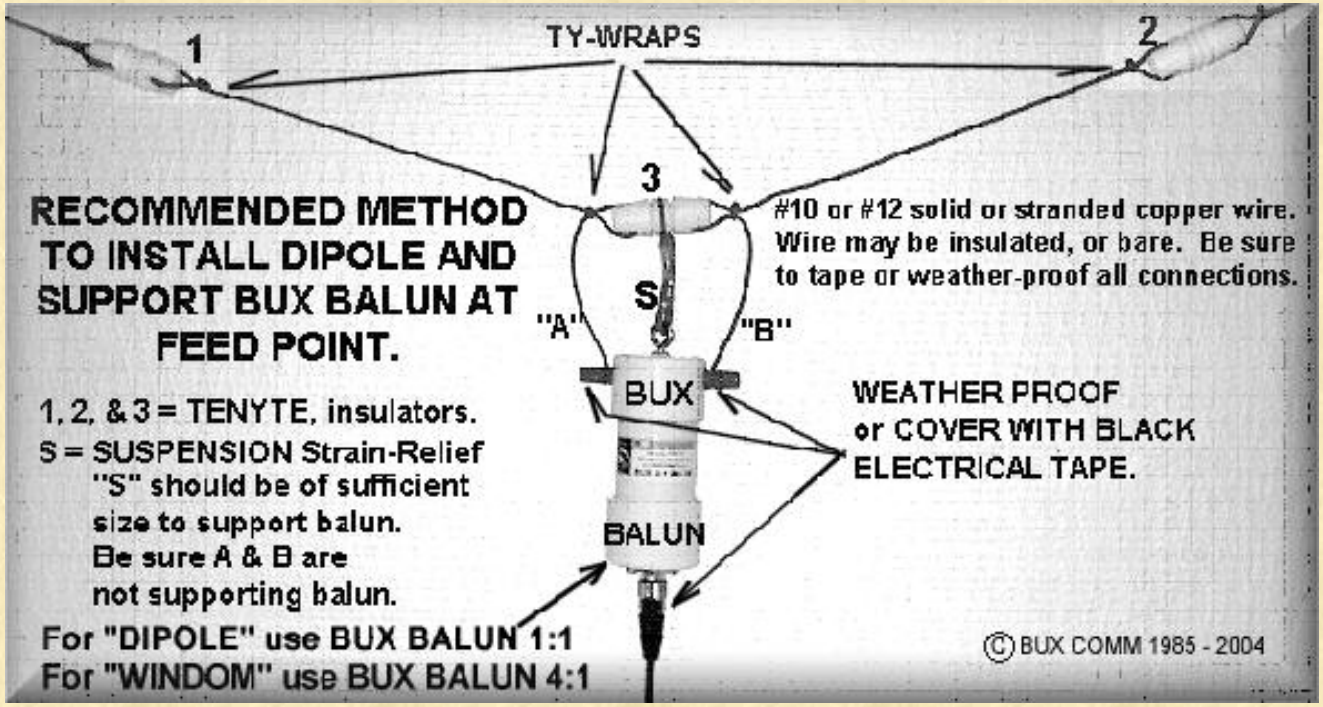


The Windom can be installed as an inverted VEE, or as a sloper, ***but in no case,*** should the angle be less than 90 degrees against itself. *To use an angle that folds against the pattern of the opposite end of the Windom, would change the impedance of the feed-point, change the multi-band features, and most important, destroy the radiation characteristics of the antenna.*

In other words, install the Windom as you would any other dipole, while using a common sense approach. The fact that we are feeding
this Windom using *coax,* and a single balun, gives us the ability to construct it as an inverted VEE or at a "real estate saving (angle)" without destroying the features and performance of the Windom.

This is *not* true with those antennas fed with ladder-lines and those that have several impedance transitions built into the feeders

of the "*basic" Windom..... There's more to come, so read on!*



ANTENNA INSULATORS

Weatherproof, (TENYTE) insulators. Perfect for your DIPOLE or Windom antenna.

For the apartment dweller, you can now hang the 20 meter doublet in the attic. I've QSO'd with stations all over the world with the 33 ft dipole in my garage attic. One insulator at the center, and one each end.

It's great for other HF WINDOM and single-band dipole antennas. Dielectric strength is comparable to the old ceramic insulators, without susceptibility to cracking or breakage under impact or extreme temperature changes.

Package of 3, $2.99 ANTINSL3 Order Now

Package of 10, TeNyte Insulators $8.99, ANTINSLX10 Order Now

Now it's time for me to P O the wire and cable vendors, and the proponents of Windom's with too many feed-line transitions. Twin Lead, Ladder-line, Window Line..... etc.. is an over-kill.. The trade-off is not...; I repeat; NOT... worth the pain and upkeep to replace it every year or two.

And trust me... It is NOT a problem of the ladder-line insulation quality or properties that I'm speaking of. I speak from experience; Wind will destroy ladder/twin-lead line. Even the robust type will succumb to the wind element in time.

IF; you decide to use ladder line, make one turn (twist) to each 20 to 24 inches of line to decrease the wind resistance presented by the "flat" line..... even if it is "window-slotted" type ladder-line. By adding the one twist per 20 inches, it may last longer than three years.

COAX does NOT present a high degree of wind resistance, and it'll last much longer.

Having *been there, done that*, and heard similar horror stories from others; another question arises regarding parallel line currents that come with the use of so-called balanced (twin-lead) feed-lines. To add injury to the ladder-line proposition, the balanced line may also assume the properties of a *single wire feeder (yuk)*.

Some purveyors of the Windom that use ladder-line transitions, must use two impedance matchers (or BALUNS) with their (knock-off) Windom. One is to transition the coax outside the HAM shack *(heaven forbid, we use ladder-line inside the shack... RF "feedback" in everything)*, to the ladder line, then another at the Windom (antenna) feed-point to choke off parallel current from the ladder-line.

Since the feed point of this antenna has been found to be near 220 to 260 ohms, *the use of a 4:1 BALUN to join our coax to the antenna,* emerges as a compromising solution.

Let's not lose sight of the most important advantage of using this antenna; and that is: It provides us with a powerful, multi-band antenna, and a minimum of feedline components.

Now for the "perfectionists" among us, if you want to smooth out the hills and valleys on some of the bands, add your antenna tuner to the system, and the Windom becomes very flat across the bands. When I say "flat," we are talking about reducing the VSWR to a minimum, to produce a good forward power, transfer of RF energy to the Windom.

We're having fun already... with the Real WINDOM antenna

The "Windom Antenna" was described by Loren G. Windom in QST magazine, September 1929. Pages 19 through 22. It is named after its inventor/designer.

Loren Windom, W8GZ, was first to reveal the antenna to the radio amateur community by describing the antenna in the September 1929 issue of QST. It was by Windom's name that the antenna became known. The Windom antenna is an off-center fed dipole with an unbalanced coax feedline.

In 1937, the Windom was first described as a compromise multiband antenna. The antenna can be employed on 80, 40, 20 and 10m with considerable, though acceptable levels of VSWR. What became perhaps the most popular multiband Windom design of all, was the German-made Fritzel FD4 antenna, described by the late Dr. Fritz Spillner1, DJ2KY, in 1971. It had the same dimensions as the multiband Windom antenna, but fitted with a 200Ω (4:1) balun in its feedpoint and fed with coax.

Today, many radio amateurs are using multiband Windom antennas with more than satisfactory results. It would not be without reason that Windom antennas are being employed during IARU HF World Championships! and most of all, by "high-stake-contests." ***Perhaps many young hams ignore the multiband Windom antenna because of its sheer simplicity and may be thinking it is too good to be true.*** *The complexity of feeding other dipoles and doublets, the losses in dipoles with traps and the esoteric marketing of some other antennas seem to appeal to them more.*

CLICK HERE: and read more about the evolution of the WINDOM, to ZEPP, to VHF J-POLE.

BUX BALUNs should be installed at the antenna feed point, or where the coax or feed-line attaches to the above ground antenna. BUX BALUNs are used to connect balanced antennas to unbalanced transmission lines, such as coax cable. Their primary purpose is to prevent antenna (RF) currents from flowing down the outside of the cable. Another function of the BUX BALUN41 is to match the impedance of an unbalanced coax to the balanced feed point of a balanced input antenna(s). BUX BALUNS may also be used as “line isolators” anywhere along the cable to prevent the destructive influence of induced RF currents (VSWR).

BUX 1:1 BALUNs are current BALUNs. They consist of several large, number 73, ferrite type 44 cores. ***BUX BALUN11*** operate from 3.5 to 72 MHz and allow use of RF power to the rated capacity of the BALUN.

BUX 4:1 BALUNs are voltage BALUNs. They consist of a large, number 41, ferrite dough-nut bobbin. ***BUX BALUN41*** operate from 3.5 to 55 MHz and allow use of RF power to the rated capacity of the BALUN.

During this month, when you purchase either of our BUX BALUNs, we will include 3 (Tenyte™), insulators FREE. A $2.99 value. *Be sure too mention this offer in*

the "Comments" space on the order form check-out page!

At BUX COMM, *We don't cut corners!

The components used in the manufacture of our BALUNs are from top quality components, beginning with the Silver Plate SO239 connectors and center insulator is made of teflon™(E.I Dupont). The wire we use to wind the ferrite donut is heavy-duty, silver flashed wire, with teflon™ insulation that will handle RF voltages above 5000 volts, and temperatures above 2000 degrees. The binding posts are heavy-duty, tempered brass, with end holes and side-thru holes to accommodate either type loop-thru connection. A double-shoulder brass capture nut is used to add a secure bite and improve antenna wire electrical connections. With our BUX UN UN (ONION), the coax is Belden™ and the PL259 connectors are Amphenol™.

4:1 Balun, BUXBALUN 41 $19.95 Order Now

o 50 ohm, SO-239 unbalanced input
o Balanced output
o 1.6 to 50 MHz
o Toroid (Voltage) design
o Heavy Duty, Lightweight construction
o Sealed against moisture

1:1 Balun, BUXBALUN 11 $19.95 Order Now

o 50 ohm, SO-239 unbalanced input
o Balanced output
o 1.6 to 50 MHz
o Toroid (Current) design
o Heavy Duty, Lightweight construction
o Sealed against moisture

BUX UN UN *De-Coupling transformer, similar to above,* but has *SO-239 (female) input* connector and *output connector is 1 ft Mini 8 cable with PL-259 (male).*

BUX ONION $19.95 Order Now

4:1, *1.5kw* Balun, BUX BALUN 41HD $27.95 Add To Cart

Toroid design, wound with teflon covered, silver plated wire.* Heavy-Duty, construction.

1:1, *1.5kw* Balun, BUX BALUN 11HD $27.95 Add To Cart

Toroid design, wound with teflon covered, silver plated wire.* Heavy-Duty, construction.

BUX ONION Decoupling Transformer

BUX UN UN De-Coupling transformer, has SO-239 (female)input connector and output connector is 1 ft Mini 8 cable with PL-259 (male).

BUX ONION is and UNbalanced to UNbalance (UN-UN) *decoupling transformer* designed to be used specifically with **the DBLSPCL antenna.**

High RF currents traveling along the coax feed-line shield can cause high VSWR. This decoupling transformer prevents RF currents from traveling down the outer shield of the coax.

The input connector is an SO239 (female) and the output connector is a PL259 (male), which mates the connector of the "DBLSPCL" RV/Apartment dweller antenna shown above.

BUX ONION $19.95 Order Now

WHY USE A 4:1 BALUN

Krusty Olde Kurt is now going to repeat himself. Why? Because the same question keeps coming up over and over. And he wants everyone to get it right.

"I'm feeding my dipole with 600-ohm line. At the station end I need a balun to convert to 50-ohm coax. I need a 12:1 balun, right?"
Wrong! *A **4:1 balun** would be better.*

Why is that? If your dipole is up, let's say, 35 feet then on 80 meters it will probably have a resistance at resonance of about 40 ohms. The actual resistance depends on the height above ground in wavelengths.

If the dipole is 40 Ohms then what do you see at the transmitter end of your 600 ohm line? If the line is a half-wave long (120 ft on 80 Meters) you'll see 40 ohms. Remember, a half-wave line repeats what it sees at the other end. But if it is a quarter-wave long you'll see 8500 Ohms! At other line lengths you'll see impedances somewhere between these two extremes.

So you are not going to see 600 ohms at the end of your 600-ohm line. That only happens if you have a 600-ohm antenna hooked onto it. With such a variation in impedance at the trans-mitter end of the line there is no one balun transformer that will match it. Most of the time the impedance will be above the 50 Ohms of your coax so a high impedance balun would be desirable. Unfortunately high imped-ance baluns don't work well when not matched.

Experience has shown that ***4:1 baluns work best*** in this service. They are more rugged and will take bad mismatches especially if they are wound on an iron powder core. So stop searching for that 12:1 balun. **Use a 4:l BALUN and your system will work just fine.**

You can read Kurt N. Sterba "AERIALS" column in **World Radio Magazine.**

BUX MOBALUN

When you hear those strong HF mobile signals, here's one reason they stand out from the rest.
By installing the BUX MoBalun near the input to your antenna, you deliver more RF energy to the antenna. At the same time, the BUX MoBalun prevents RF from traveling back along the shield (high SWR) of your coax. High power rating, Low-permeability toroid, with Internal composition fiber-glass to prevent vibration during mobile operation. For input and output connectors, we use only the best Silver plated, Teflon insulated, SO239 connectors. 700 watts PEP. *Our **MoBALUN** is also ideal for marine antenna installations*.

$19.95 MOBALUN Order Now

ANTENNA INSULATORS

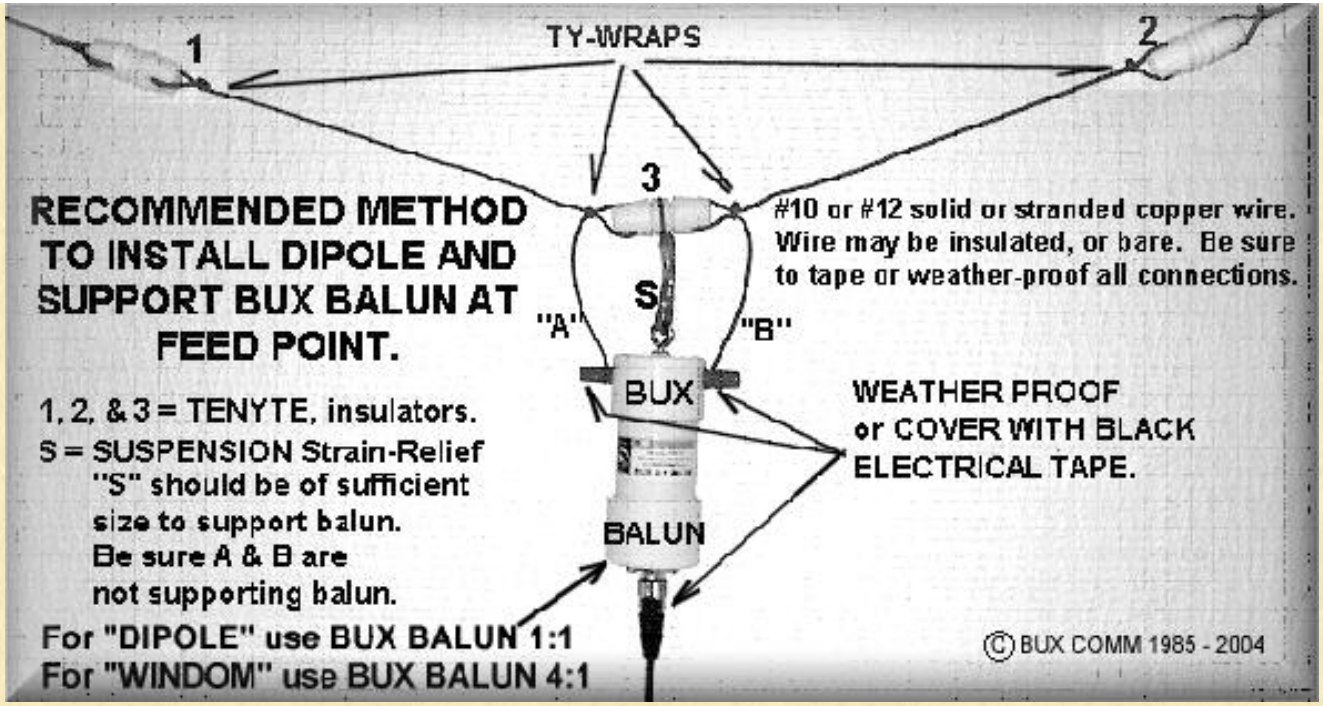
Weatherproof, (TENYTE) insulators. Perfect for your DIPOLE or Windom antenna.

For the apartment dweller, you can now hang the 20 meter doublet in the attic. I've QSO'd with stations all over the world with the 33 ft dipole in my garage attic. One insulator at the center, and one each end.

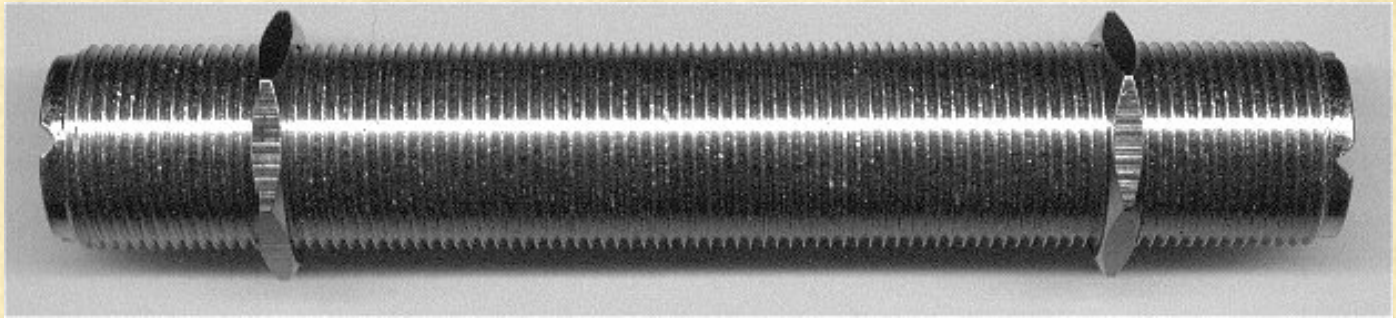
It's great for other HF WINDOM and single-band dipole antennas. Dielectric strength is comparable to the old ceramic insulators, without susceptibility to cracking or breakage under impact or extreme temperature changes.

Package of 3, $2.99 ANTINSL3 Order Now

Package of 10, Insulators $8.99, ANTINSLX10 Order Now



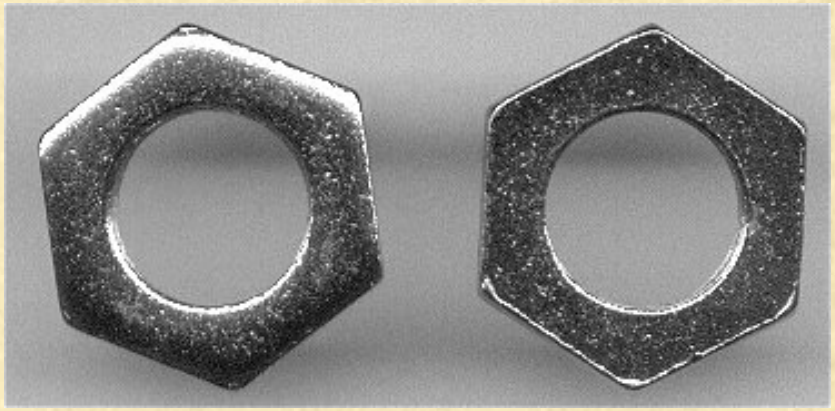
UHF DOUBLE FEMALE BULKHEAD (feed-thru) CONNECTORS



For bulkhead and through-the-wall UHF connector feed-thru connections, with keeper nuts.:

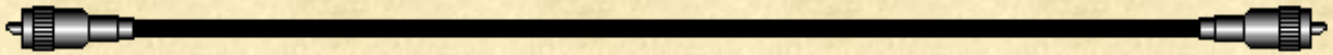
Order 7518-2 (Two inches)	**$ 2.95 ea.**	Order Now	**10 for $ 24.50**	Order Now
Order 7518-4 (Four inches)	**$ 3.95 ea.**	Order Now	**10 for $ 34.50**	Order Now

Order 7518-6 (Six inches)	**$ 4.95 ea.**	Order Now	**10 for $ 44.50**	Order Now
Order 7518-8 (Eight inches)	**$ 8.95 ea.**	Order Now	**10 for $ 82.50**	Order Now
Order 7518-10 (Ten inches)	**$ 10.95ea.**	Order Now	**10 for $ 99.50**	Order Now
Order 7518-12 (Twelve inches)	**$12.95ea.**	Order Now	**10 for $ 114.50**	Order Now



Heavy Duty (1") Nuts for the above bulkhead connectors.

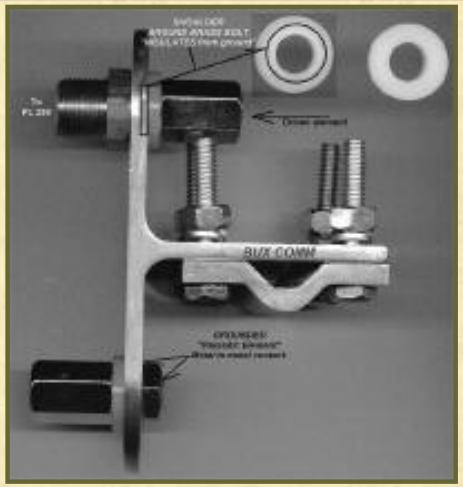
2/.99 cents HDM1-NUT Order Now **10 for $ 3.99** Order Now



50 Ohm impedance, 3 ft, 6 ft, 9 ft, 12 ft, & 18 feet.

These jumpers are made from high quality, low-loss, coax cable with PL-259 connectors installed at each end.

Three ft. (3')	**ORDER No. 8X3-PLPL $3.95**	Order Now	
Six ft. (6')	**ORDER No. 8X6-PLPL $4.95**	Order Now	
Nine ft. (9')	**ORDER No. 8X9-PLPL $5.95**	Order Now	
Twelve ft. (12')	**ORDER No. 8X12-PLPL $7.95**	Order Now	
Eighteen ft. (18')	**ORDER No. 8X18-PLPL $8.95**	Order Now	



Reinforced, Double Bracket used with our "DBLSPCL" antennas shown elsewhere on this page.

Mount two mobile whips in a horizontal plane to form a compact apartment dweller Dipole, or RV antenna. Comes with two sets of mounting hardware (bolts), those shown here and another set of longer bolts for larger masts. Double Antenna Bracket as shown at left.

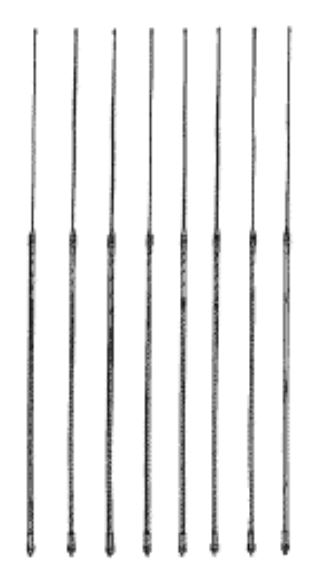
$16.95 DBLBRKT

Add to Your SHOPPING CART

After a few years, the hardened, nylon shoulder washers become weather-brittle and break. Keep a few spares in the Ham Shack:

4 for 99¢ SM1SW

Add to Your SHOPPING CART



Mobile *tuneable-tip* antennas.

BHF75M 75 METER 3/8" X 24 THREAD 8 feet long $16.95

Add to Your SHOPPING CART

Here's the perfect solution for your RV and portable HF antenna applications.

Use our dipole, mast-mounted bracket for 'fixed-station', field-day operating, apartment dwellers, and the RV travelers. With two of the mobile HF antennas shown at left, you can have the best of both worlds.

BHF40M 40 METER 3/8" X 24 THREAD 8 feet long $16.95 Add to Your SHOPPING CART

BHF20M 20 METER 3/8" X 24 THREAD 8 feet long $16.95 Add to Your SHOPPING CART

BHF17M 17 METER 3/8" X 24 THREAD 8 feet long $16.95 Add to Your SHOPPING CART

BHF15M 15 METER 3/8" X 24 THREAD 8 feet long $16.95 Add to Your SHOPPING CART

BHF10M 10 METER 3/8" X 24 THREAD 8 feet long $16.95 Add to Your SHOPPING CART

BHF6M 6 METER 3/8" X 24 THREAD 5 feet long $16.95 Add to Your SHOPPING CART

Spare 48" Tips for above *tuneable-tip* antennas.TIP48, $2.95 Add to Your SHOPPING CART

By combining a pair (two) of the single band HF Mobile (loaded) tuneable-tip antennas, and (VHF) whips at left, you can have an ideal, single element, horizontal dipole (beam). This type antenna opens the world of HF and VHF communications for the apartment dweller, RV HAM, field-day operating, and a variety of other uses and applications. Shown mounted on 1 & 1/4" mast. Antenna(s) Not Included.

$16.95 each

DBLBRKT Add to Your SHOPPING CART

tuneable-tip antennas allow adjusting for minimum VSWR.

CLICK HERE to read more about our HF DIPOLE SPECIAL!

By combining a pair (two) of the single band HF Mobile (loaded) tuneable-tip antennas above, you can have an ideal, single element, horizontal dipole (beam). This type antenna opens the world of HF communications for the *apartment dweller, RV HAM, field-day operating,* and a variety of other uses and applications. The packages described below include two antenna, the Double Antenna-Bracket (DBLBRKT), and mounting hardware

$ 39.97 DBLSPCL (without BALUN) *With BALUN, see "DBLCOMBO", below.*

75 Meters→ Add to Your SHOPPING CART, 40 Meters→ Add to Your SHOPPING CART, 20 Meters→ Add to Your SHOPPING CART,

17 Meters→ Add to Your SHOPPING CART, 15 Meters→ Add to Your SHOPPING CART, 10 Meters→ Add to Your SHOPPING CART,

6 Meters → Add to Your SHOPPING CART

By combining a pair (two) of the single band HF Mobile (loaded) tuneable-tip antennas above, you can have an ideal, single element, horizontal dipole (beam). This type antenna opens the world of HF communications for the *apartment dweller, RV HAM, field-day operating,* and a variety of other uses and applications. The package described below includes two antenna, the Double-Bracket (DBLBRKT), mounting hardware, and BUX BALUN (UNUN). $AVE *$$$$$*, ADD the BUX UNUN , shown below, BOTH, DPLSPCL & BUXUNUN, *for only $ 54.97*, *ORDER, DBLCOMBO*

75 Meters→ Add to Your SHOPPING CART, 40 Meters→ Add to Your SHOPPING CART, 20 Meters→ Add to Your SHOPPING CART,

17 Meters→ Add to Your SHOPPING CART, 15 Meters→ Add to Your SHOPPING CART, 10 Meters→ Add to Your SHOPPING CART

6 Meters → Add to Your SHOPPING CART

FAX 434 525 7818 or Mail order form, Click Here!

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EVINGTON, VIRGINIA 24550

FAX 434 525 7818

6m Vertical Antenna

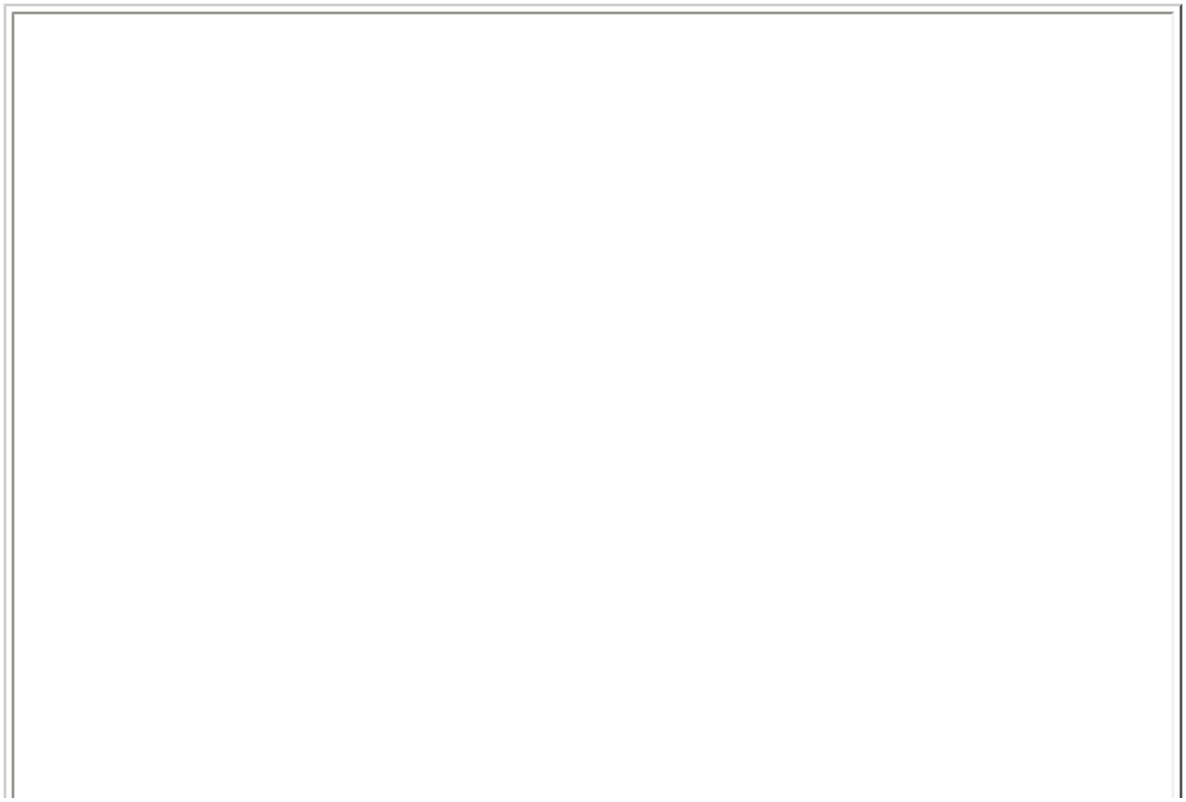
By Mike, G3JVL

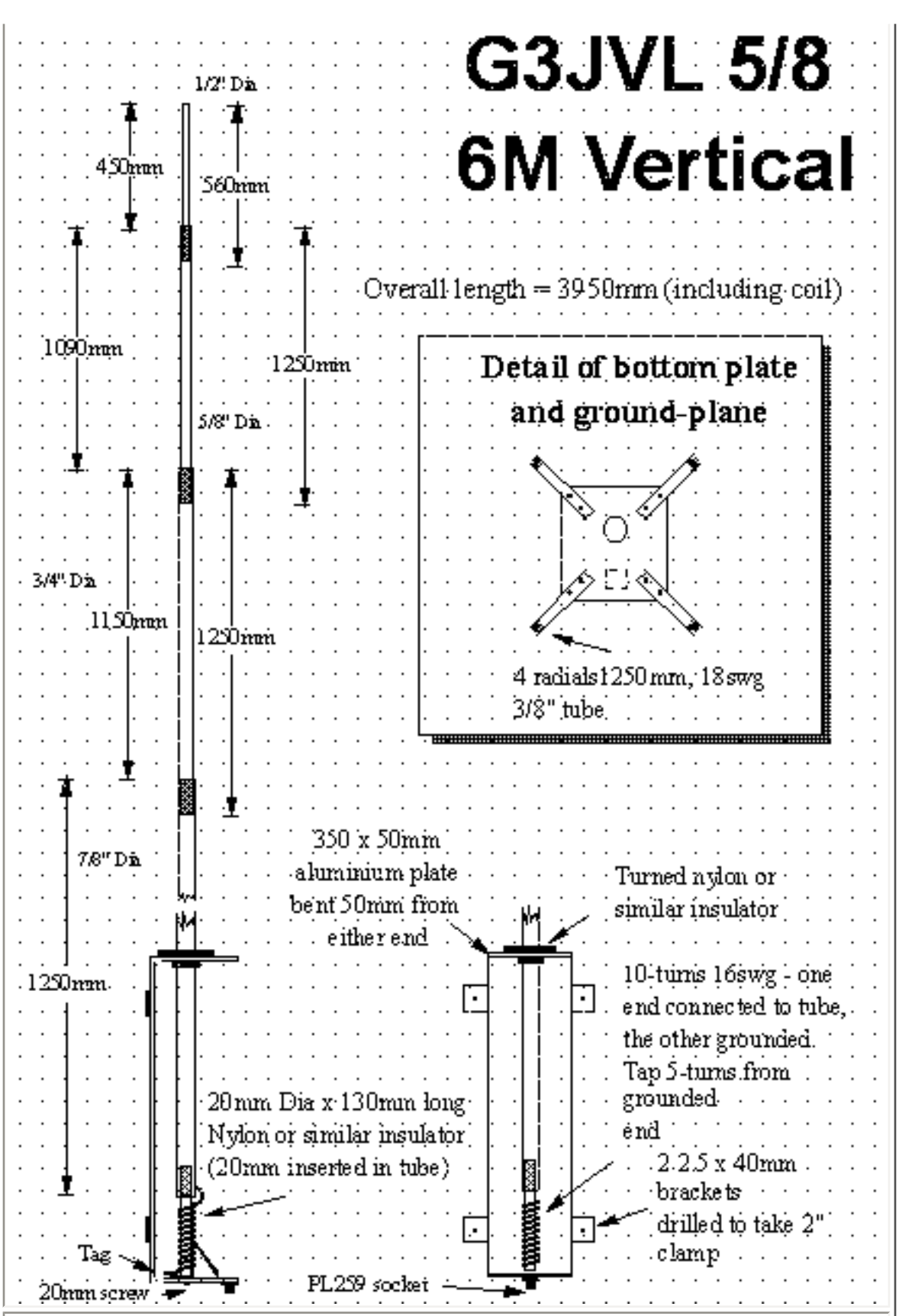
The G3JVL 6M ground plane vertical is a compact antenna that is ideal for portable operations. It packs away into a small bag only 1.3 metres long which is an ideal size for hand-baggage on aircraft.

Although a little engineering work is required, it well within the abilities of the home workshop and all you newly licensed amateurs!

The vertical itself is constructed out of four overlapping sections of aluminium tube whose sizes are given in the diagram opposite. The four tubes are meant to telescope so wall thickness should be chosen to achieve this. In practice, 16swg or 1.6mm might be OK but 18swg or 1.2mm will easily fit. The lengths are held in place by three stainless steel self-tapping screws.

The vertical is bottom-loaded with a coil wound on an insulated former (nylon or similar material - it is not too critical at 50MHz). - the former is 130mm long with the top 20mm turned so it can be inserted into the bottom section of the vertical. The loading coil consists of ten turns of 16swg (1.6mm) diameter copper wire with one end connected to the bottom section of the vertical with a machine screw and the other end connected to the ground plate. A tap at 5-turns is connected to the input PL-259 socket.





The construction of the 6m vertical antenna

The four ground plane elements are Constructed from 1250mm long, 9.5mm (3/8") diameter aluminium tube. These are mounted to the base plate by the use of eight

stainless steel machine screws. This arrangement allows easy disassembly.

The vertical itself is supported by a piece of 350mm by 50mm x 4mm aluminium plate bent into a 'U' shape. The bottom section of the vertical is insulated by a turned piece of nylon or similar tubing. If you do not have access to a lathe any other insulating arrangement should suffice so long as it is robust.

The overall length of the vertical is 3950mm including the loading coil. The top section should be trimmed to set the centre frequency. If the VSWR at resonance is not close to 1:1 then alter the position of the tap on the loading coil (remember, changing this will alter the resonance of the antenna!).

Once completed give the whole assembly several coats of varnish to keep out the weather.

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The ON4ANT antenna page

An overview of interesting antenna designs and links to antenna sites.

Contents

- HF designs
- 6m designs
- VHF designs
- Commercial designs and links to designer software
- Antenna pictures
- My personal 6m site

HF designs.

Several sites on the web carry very good information about antenna designing. One of the most interesting sites I've found is that of W4RNL, have a look t them. In the past i made a 5 band yagi, covering 20 to 10 meter band, with a boom length of 40-50-60ft. This design was published in QEX a while ago. It has been reproduced successfully

by several hams around the world. This killer antenna was in use for our OT9E OT0E

Contest station. We managed to win 3 times in a row the ARRL roundup as multi single.

Contents.

- A multiband yagi covering 10 to 20 meters.
- A 5el 10m beam (wideband design)
- A 5el 10m beam on a 6m boom

Back to top

6m designs

An overview of some really big yagis for 6 meters. For those with limited space there are some

short yagi designs available too.

Contents.

- 9 element yagi on a 15 meter boom
- 9 element yagi on a 12 meter boom using a folded dipole

Back to top

VHF designs

- under construction

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Commercial designs and links to designer software

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A multiband yagi antenna

Design en construction ON4ANT-ON4GG

As most of us know, the monoband yagi is by far the best antenna choice. The majority of hams have unfortunately no room to put up several towers for all the different monobanders. The average ham chooses a trapped multi-band yagi. This antenna type allows him to be active on a number of bands, but it has some drawbacks as well: loss of swr bandwidth, antennagain and F/B ratio. Over the past years a number of commercial interlaced designs have been available. These designs often put 2 bands on a same boom.

These interlaced yagis often give a good result and can be an excellent replacement for the trapped yagis.

(The W4RNL web site carries an interesting article about these interlaced yagis).

Struggling to get a number of bands with good swr bandwidth and gain on a single boom made me decide to develop the antenna described here. The basic principle is to put a number of mono band yagis on the same boom, one in front of the other. The first conclusion is that the boom length increases rapidly, especially if one wants to cover 20 to 10 meters.The boom length was limited to 15 meters with an option to shorten the boom to 12.8 meters. This should allow most of us to reproduce the design. Those having plenty of room can go for the long design 18.3m (60ft) boom. The antenna covers the 20 to 10-meter bands.

The design has been done with the help of AO*, YO*, EZNEC/4*, STRESS*, and YAGI DESIGN*.

The electrical design can be found in part 1.

It gives full details about element lengths and spacing. The feedpoint impedance, free space gain and swr bandwidths are also given. Two modified designs are described as well.

Part 2 gives mechanical details, including the tapering detail, wind survival and total wind load.

Part 3 gives a table with the gamma match details

Part 4 gives you a table with gain figures from the most common available mono band yagis. These designs have been verified with the same software as used for designing this antenna. Trapped yagis have not been taken into account as these show less gain than their mono band counterparts and usually these trapped yagis have unrealistic gain figure claims.

-

Part 1.

The basic calculation has been done for an antenna in free-space and all values are in dBi. We don't take into account the influence of the earth ground gain, and the reference antenna is an isotropic radiator. (0 dBd = 2.15 dBi). If one takes into account the ground gain (as most manufacturers do) the gain figures will be 4-5 dB higher.

However this in influenced by the antenna height.

The setup above real ground will change the radiation pattern.

The table gives the element length for a constant diameter (20mm) and the element spacing.

Element Length	Description	Position (m)
5.45	Reflector 20	00.00
5.2	Driver 20	02.00
4.9	Director 20	03.60
4.15	Reflector 17 & dir 20	05.25
4.02	Driver 17	06.20
3.8	Reflector 15 & dir 17	07.20

3.395	Driver 15	08.40
3.02	Director 15 & ref 12	09.50
2.91	Driver 12	10.30
2.78	Reflector 10 & dir 12	11.60
2.55	Driver 10	12.45
2.355	Director 10	13.40
2.265	Director 10	15.00

What is to expect from this antenna?

Gain is comparable to a 3-4-element monobander, with excellent swr bandwidth and F/B.

-

Antenna specifications.

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Frequency Gain (dBi) Impedance F/B SWR

14.000 8.1 33.0-j4.1 26.8 1.26

14.175 8.2 30.9+j3.0 29.1 1.00

14.350 8.3 26.0+j12.2 25.9 1.44

18.068 8.1 20.9-j3.6 21.5 1.10

18.118 8.6 22.3-j2.3 22.3 1.00

18.168 8.6 23.5-j1.2 23.2 1.07

21.000 8.4 32.4-j7.8 21.1 1.27

21.200 8.5 34.2+j0.5 21.0 1.00

21.400 8.6 35.7+j8.1 20.9 1.25

```
24.880 8.5 10.7-j3.6 30.6 1.19

24.940 8.5 10.8-j1.7 30.6 1.00

24.990 8.5 10.8+j0.1 28.0 1.19

28.000 7.9 26.0-j7.2 29.7 1.47

28.350 8.1 26.9+j3.1 25.7 1.00

28.700 8.2 27.6+j13.9 22.6 1.48
```

This design has an almost constant gain over the 5 bands.

The swr bandwidth is excellent over the entire range with exception of 10 meters;

here it is limited to 28.8 MHz. Of course this swr is in reference to the matching frequency.

I'm sure that things still can be improved, but this may have a negative influence on swr bandwidth and/or F/B. Another disadvantage of getting the last .5 dB out of the design makes it more critical and less tolerant for small dimension errors (element lengths and spacing).

If you really want more gain, go for the longer design on the 18m boom. You will get the same bandwidth and F/B (or even better) with higher gains.

Variant 1.

A 15-meter boom too big for you? Perhaps this 12.8m antenna is the solution.

There will be one element less on 20m. The gain will drop to about 7 dBi, which is still good.

Element length (m) Description Position (m)

5.45 Reflector 20 0.00

5.2 Driver 20 2.00

4.15 Reflector 17 and director 20 3.05

4.02 Driver 17 4.00

3.8 Reflector 15 and director 17 5.00

3.395 Driver 15 6.20

3.02 Director 15 and reflector 12 7.30

2.91 Driver 12 8.10

2.78 Reflector 10 and director 12 9.40

2.55 Driver 10 10.25

2.355 Director 10 11.20

2.265 Director 10 12.80

Only 20m changes, they other gain figures remain.

Antenna specifications.

Frequency Gain (dBi) Impedance F/B SWR

_

14.000 7.2 33.5-j11.6 16.0 1.40

14.175 7.1 39.8-j0.9 29.1 1.00

14.350 7.0 45.3+j9.0 14.3 1.30

Variant 2.

Do you have plenty of room? This 18.3m monster is the solution.

It gives you higher gain on the top 3 bands with an excellent bandwidth.

Element length (m) Description Position (m)

5.45 Reflector 20 0.00

5.2 Driver 20 2.00

4.9 Director 20 3.60

4.15 Reflector 17 and director 20 5.25

4.02 Driver 17 6.40

3.8 Reflector 15 and director 17 7.20

3.395 Driver 15 8.40

3.02 Director 15 and reflector 12 9.50

2.91 Driver 12 10.80

2.68 Reflector 10 and director 12 12.00

2.55 Driver 10 13.014

2.47 Director 10 13.816

2.44 Director 10 15.775

2.31 Director 10 18.25

_

Antenna specifications.

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Frequency Gain (dBi) F/B

_

14.175 8.3 34

18.118 8.3 21

21.200 8.7 23

24.940 9.6 38

28.350 10.0 29

This design made it at my home QTH.

The calculated specification seem to be corresponding really well with the on air performance.

Initial testing show an advantage as compared to a very large commercial multi band yagi.

The design is very broadband and allows different kinds of matching.

_

Part 2.

_

Feeding the antenna.

The driven elements are all resonated in band.

The actual impedance of the antenna is high enough to allow different kind of feeding.

Personally I use a gamma match; the elements don't need to be spliced up in this case.

Element mounting.

One can choose isolated or non-isolated element mounting. The boom influence on the element length is minimal.

The use of isolated element can be a disadvantage is you want to use your tower as a toploaded vertical on 160m. (* ON4UN).

The boom element plate measures 200x100mm. If you wish to mount the elements non-isolated you can calculate the

influence of the boom on the element lengths with **YAGI DESIGN***.

The calculated influence is only a few millimeters for the 20m element. As this design is not critical, one can use the isolated element lengths.

Element tapering.

Each element has to be as strong as possible for a minimal windload and weight so we use tapering.

Most of the available antenna design software programs allow calculating the taper. Only a few allow calculating the element strength.

Initially I used **STRESS***, this software is used by the former Telex/Hygain company.

Afterwards I used a Belgian product, **YAGI-DESIGN*** by ON4UN. This package can calculate in all circumstances the taper of an

element that complies with a wind survival. This for the lowest possible weight and windload. The element sag is also calculated.

The calculated minimal wind survival is 160km/h, (100mph). The antenna is mounted on an 80ft tower on top of a 300ft hill here.

Parameters: EIA-222-C pressure 30lb/sq ft at 86mph.

Shape factor .666

No ice-load

Aluminum 6061-T6 (yield strength 35000)

The table gives us element diameter, wall thickness, length, half element weight and length. The elements will be adjusted with the tip end.

Some of these elements are telescopic on the inside. All of the 20m element consist of 3 diameters.

–

Element 1-2-3

```
1700x28x1.5 3000x25x2.5 2750x19x1.5
```

Insert 250x19x1.9

The wind load of this half element is 0.13m². The weight of this half element is 2.6kg

Element sag is 20.5cm. The tip will be adjusted.

Tip lenght ! (100mm overlap)

Isolated Non-isolated

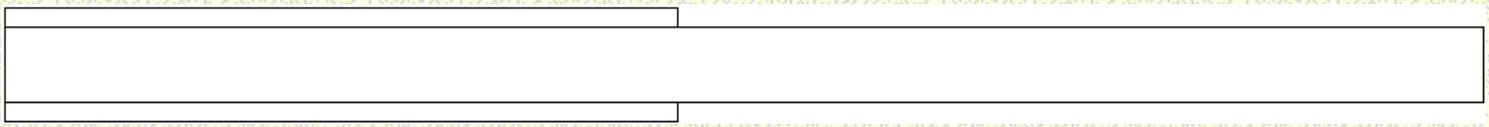
Element 1 : 2567mm 2570mm

Element 2 : 2309mm 2312mm

Element 3 : 1998mm 2000mm

Element 4-5-6

1750x25x2.5 **x19x1.5



The wind load of this half element is 0.084m². The weight of this half element is 1.85kg

Element sag is 8.4cm. The tip will be adjusted

Total element length !

Isolated Non isolated

Element 4 : 4222mm 4224mm

Element 5 : 4053mm 4056mm

Element 6 : 3852mm 3854mm

Element 7-8-9

_

1500x25x2.5 ****x19x1.5

The wind load of this half element is 0.074m². The weight of this half element is 1.2kg

Element sag is 6cm. The tip will be adjusted

Total element lenght ! (100mm overlap)

Element 7 : 3450mm

Element 8 : 3082mm

Element 9 : 2957mm

Element 10-11-12-13

750x25x2.5 **x19x1.5

The wind load of this half element is 0.059m². The weight of this half element is 1.1kg

Element sag is 3.5cm. The tip will be adjusted

Total element length ! (100mm overlap)

Element 10 : 2845mm

Element 11 : 2583mm

Element 12 : 2364mm

Element 13 : 2264mm

-

Wind load and weight of elements.

44.55 kg and a 2.20 m² windload.

If you choose the variant 1 you will have 0.26m² less wind load and will gain about 5.2 kg.

The actual weight of the antenna is function of the choosen boom diameter, the mounting plates and all related hardware.

My antenna uses a 4 inch boom and the weight is around 60 kg.

Part 3.

The Gamma match for the 18.3 m version

	A(mm)	B(mm)	C(pF)	D(mm)	D(mm)
14.175	1067	120	180	28	4
18.118	562	120	150	25	4
21.200	735	100	100	25	4
24.930	867	100	100	25	4
28.400	361	100	81*	25	4

* (3x27pf parallel)

The Gamma match for the 15 m version

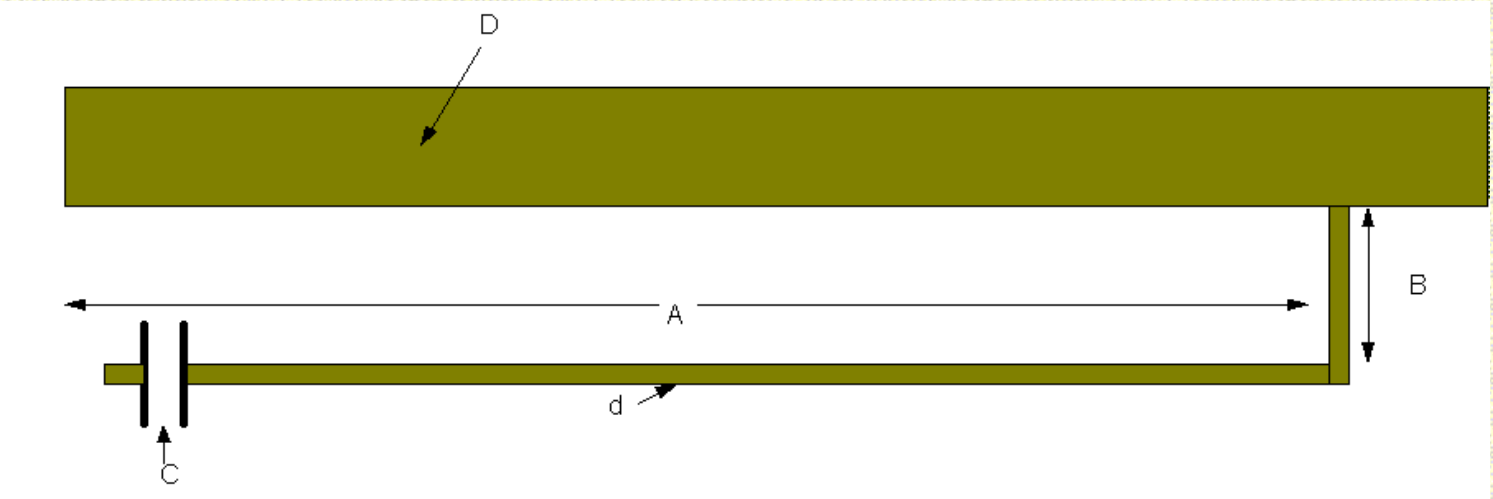
	A(mm)	B(mm)	C(pF)	D(mm)	D(mm)
14.175	950	120	180	28	4
18.118	602	120	180	25	4
21.200	631	110	100	25	4
24.930	522	80	270	25	**3**

28.400	496	100	100	25	4

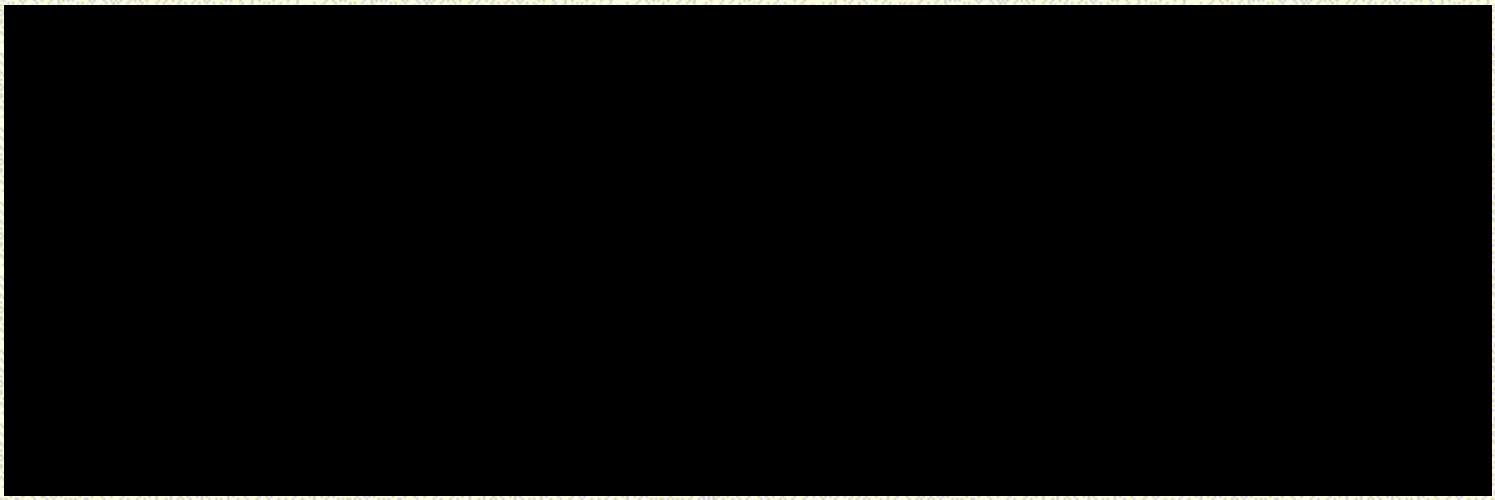
[The Gamma match for the 12.8 m version]

	A(mm)	B(mm)	C(pF)	D(mm)	D(mm)
14.175	880	120	154	28	4
18.118	598	120	200	25	4
21.200	629	100	111	25	4
24.930	279	100	190	25	**3**
28.400	497	100	100	25	4

For an output power of 2000 watts one should get capacitors handling 300 volts and 6.3 amps current.



[The Hairpin match for the 18 m version]



	A(mm)	B(mm)	d(mm)	Shorten driven element
14.175	50	541	4	139mm
18.118	50	393	4	56mm
21.200	50	407	4	88mm
24.940	50	235	4	87mm
28.500	50	317	4	103mm

Part 4.

Is it all worth the trouble ? Looking at the actual cost, YES. The price should be below $800 (US) for the 60ft design.

The design is non-critical can can be easely reproduced. The gain is excellent and you will have a big signal on the bands.

However, an antenna this size requires a strong tower and big rotator. If you have the tower and rotator for it, its an excellent choice.

A comparison with some commercial mono band antennas gives you a idea about the performance of this antenna.

The values indicated are NOT those from the manufacturer, but those calculated with the design software used for this antenna.

Only this procedure gives an objective view on the gain, since all gains were computed in exactly the same way.

dbi

Type Gain SWR bandedge Description

310-08 7.17 1.31-1.46 3el 10m on 2.3m boom

103BA 7.51 1.53-2.01 3el 10m on 2.3m boom

153BA 7.68 1.45-1.68 3el 15m on 3.5m boom

315-12 7.54 1.49-1.45 3el 15m on 3.6m boom

320-16 7.21 1.27-1.38 3el 20m on 4.7m boom

203BA 7.17 1.22-1.20 3el 20m on 4.8m boom

20-3CD 8.09 2.03-2.90 3el 20m on 6.0m boom

10-4CD 8.58 1.63-1.79 4el 10m on 4.8m boom

412-15 8.40 1.09-1.09 4el 12m on 4.4m boom

415-18 8.24 1.41-1.38 4el 15m on 5.4m boom

417-20 8.52 1.08-1.11 4el 17m on 6.0m boom

204CA 8.25 1.49-1.47 4el 20m on 7.8m boom

420-26 8.60 1.28-1.37 4el 20m on 7.8m boom

20-4CD 8.54 1.78-2.20 4el 20m on 9.6m boom

510-20 9.75 1.49-1.53 5el 10m on 6.0m boom

KLM510 9.21 1.33-1.43 5el 10m on 6.1m boom

105CA 8.38 1.42-1.23 5el 10m on 7.2m boom

155CA 9.70 1.49-1.62 5el 15m on 7.7m boom

205CA 9.23 1.43-1.96 5el 20m on 10.4m boom

KLM520 9.43 1.66-1.25 5el 20m on 12.8m boom

Conclusion.

This design is a valuable alternative for a 4 element monoband yagi, taking into account the gain and swr bandwidth.

It is obvious that some improvements can be done, depending on your specific needs. Perhaps you need less bandwidth.

I tried to have a broadband yagi with gain figures close too or better than the common 4 element moband yagis.

The real gain, with associated radiation angle is given in next table.

14.150 13.55 dBi @ 12°

18.118 13.64 dBi @ 10°

21.200 13.74 dBi @ 8°

24.940 14.20 dBi @ 7°

28.400 13.77 dBi @ 6°

If you wish more info on this design, or want to share some of your antenna experiences, you can always email me ON4ANT@hotmail.com

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References

AO and YO written by K6STI.

Comparison of this design made by W4RNL with EZNEC4 shows very similar results to those obtained with AO.

The W4RNL site contains lots of valuable antenna information and really is worthwhile visiting.

My sincere thanks to L.B. Cebik for the verification of this design and the information available on his website.

YAGI DESIGN was written by ON4UN and covers all mechanical issues of antennas. It's a DOS based program,

and is extremely easy to use. I wish to thank John for his help as well. Those wishing to obtain this program can

always contact John, ON4UN

STRESS Hygain/Telex.

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A 5 element yagi for 10 meters

(Wide band design)

Here is a very simple 5 element covering the entire 10 meter band (28.000-29.700) with a swr of less than 1.4 at the band edges. The elements are non tapered! Element diameter is 11 mm.

	Position (mm)	Half element (mm)
Reflector	0	2670
Driven	1014	2530
Director 1	1816	2400
Director 2	3775	2370
Director 3	6250	2250

Here is the overview of freespace gain compared to an isotropic source. SWR reference is 28.500 MHz.

Freq.	Gain(dBi)	F/B	Impedance	SWR
28.000	9.3	16.8	31.3-j9.0	1.35
28.500	9.5	21.1	37.4-j0.5	1.00

29.000	9.8	21.5	38.7+j7.8	1.25
29.700	9.8	23.0	28.6+j3.3	1.35

Seen the high impedance, several matching methods are possible.

What happens if we use 20mm elements?

Freq.	Gain(dBi)	F/B	Impedance	SWR
28.000	9.5	17.9	31.3-j6.1	1.26
28.500	9.8	22.3	34.1+j0.7	1.00
29.000	10.0	22.5	36.4+j9.2	1.28
29.700	9.5	25.3	10.2+j9.1	3.56

Bandwidth has been reduced!! So probably you will need to re-dimension the yagi.

Lets consider a boom-element plate of 100x200mm. On the version with the 20mm element..

We start of with a saddle height of 10mm, element is mounted 10mm above the plate with U bolts).

The equivalent element diameter for the part above the plate (a length of 100mm that is) has an

equivalent element diameter of 46.4 mm. So when modelling the yagi with the plate, one has to consider

The element as being tapered, a 100mm long 46mm element in the middle and the rest a

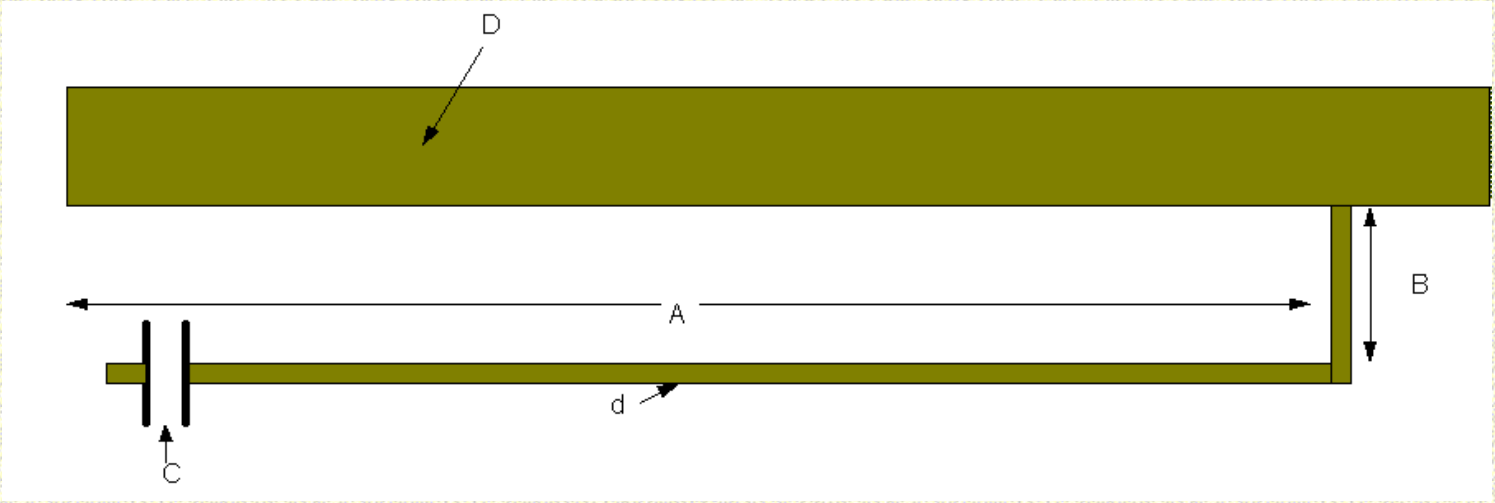
20mm element.

The boom diameter itself has no more influence on the equivalent element diameter. (if insulated)

Now we are going to verify the influence of the saddle height on the equivalent element diameter!

Saddle height	Equivalent element diameter
5mm	42.4mm
7mm	44.0mm
9mm	45.6mm
10mm	46.4mm

Gamma matching.



d = 4 mm

D = 20 mm

A = 420mm

C = 100 pf

B = 100 mm

Hairpin matching.

If a hairpin is used the driven element needs to be shortened by 10.5 cm !

A = 50 mm

B = 360 mm

D = 4 mm diameter

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A 5 element yagi for 10 meters

On a 6m boom. (standard length)

This economical design uses a standard 6 meter long square boom (50x50x1.5mm).

The boom-element plate is a 5mm thick 200x100mm plated bolted to the boom.

The elements are bolted to this plate. (no saddles)

Matching is achived with a hairpin, so the driven element is insolated from the plate.

A rubber water tube with an inner diameter of 20mm will do fine, and use 2 u bolts to fix this element. The element diameter is 20x1mm

	Position (mm)	Half element lenghts (mm)
Reflector	0000	2670
Driven	1000	2459*
Director 1	1750	2420
Director 2	3460	2380
Director 3	5950	2190

* the element is shortened to accomodate the hairpin (was 2530mm before)

	28.000	28.500	29.000	29.400
Gain	9.31	9.5	9.6	9.5
F/B	20.4	22.7	22.7	22.9
Imp.	26.9-j7.9	31.2+j1.0	36.5+j6.0	22.8-j0.8
SWR	1.34	1.07	1.3	1.3

A = 50 mm

B = 271 mm

D = 4 mm diameter

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A 9 element yagi on a 15m boom

(a 50 MHz yagi with 15.0 dBi gain)

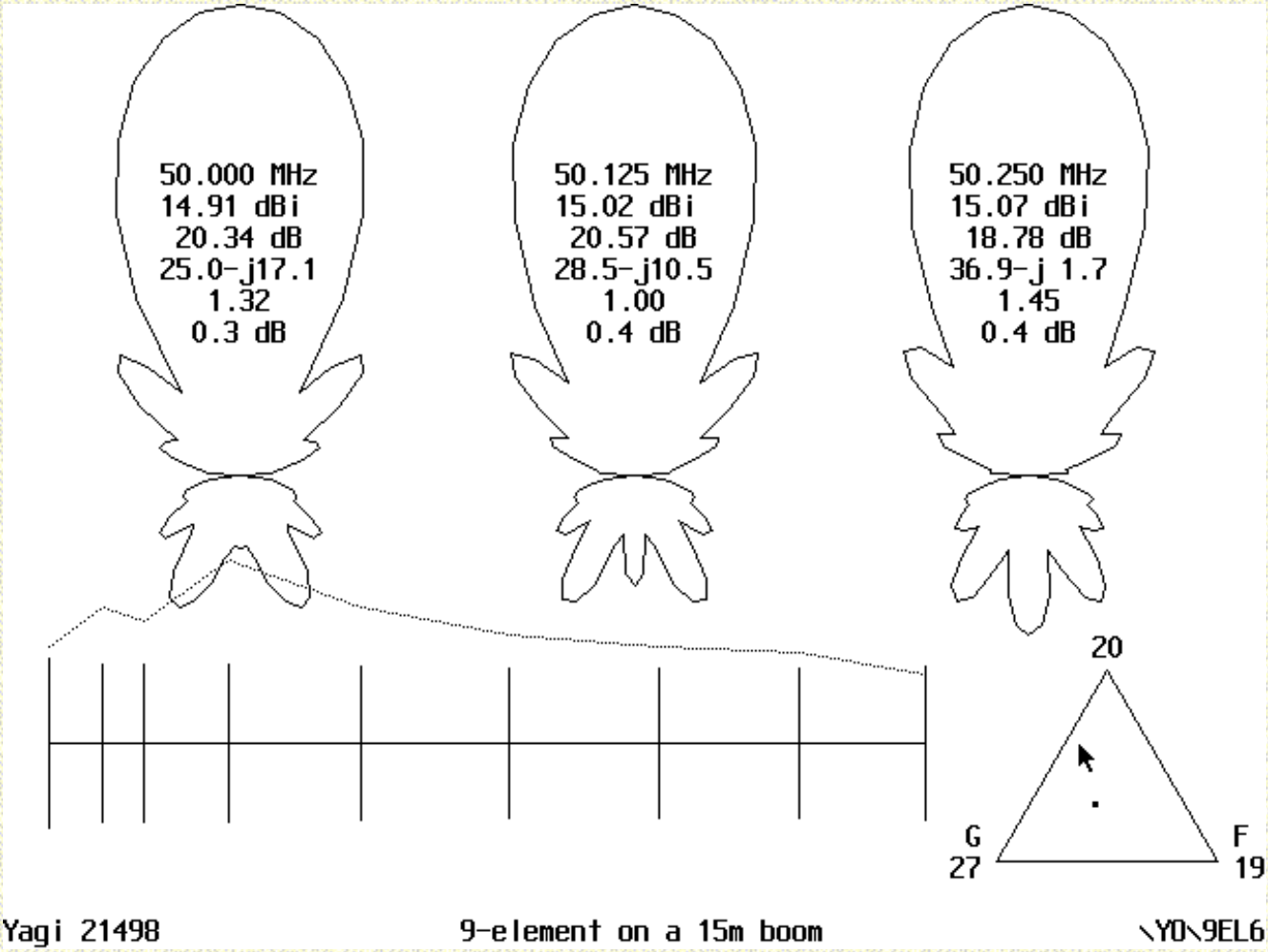
Having lots of space for antenna I started out with a design for a long yagi, 9 elements on a 15 meter boom.

I did not choose to go for a folded dipole, but rather stick to a minimum value for the feed impedance.

Table : Element length versus position. Element diameter is 12 mm.

Position (mm)	Length (mm)
0.0000	144.9851
91.5139	135.0005
163.0316	135.4751
163.0316	133.1443
531.7224	130.1802
780.1813	128.3098
1037.0823	127.7031
1273.9320	128.1430
1489.3116	130.1693

Table : Gain, F/B, SWR and impedance versus frequency.



A total gain of 15.0 dBi, this out performs the DL6WU designs by a few tenths of a dB.

It outperforms the commercial M2 design as well with 0.15dB.

(6M9KHW !, the older 9 element, a 2 WL, is 1 dB down)

The impedance remains reasonable and allows for different feed methods.

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A 9 element yagi on a 12m boom

(a 50 MHz yagi with 14.1 dBi gain)

If one wants a very simple feed like a folded dipole with a 4:1 coaxial balun, you can opt for for this design.

Lower gain (14.1dBi) but excellent match.

Table : Element position versus element length

Position(mm)	Half element length
0	**1450**
1400	**1420**
1966	**1368**
3170	**1350**
4700	**1320**
6450	**1320**
8510	**1318**
10640	**1298**
12840	**1298**

This very 'easy' design has a free space gain of 14.1 dBi. The element diameter is 12mm.

The values given in the table do not take into account the influence of the boom.

There are some formulas to calculate the influence of the boom on isolated elements going through it.

Have a look at http://www.dubus.org for valuable info on antenna designs for 6-2-70-

23.

Table : Gain, F/B, SWR and impedance versus frequency.

	50.000	50.150	50.300	50.450
Gain (dBi)	14.04	14.10	14.12	14.08
F/B (dB)	20.32	20.84	20.09	17.01
Impedance	43.6-j2.2	49.3+j0.5	57.9-j0.8	65.9-j11
SWR	1.16	1.01	1.16	1.28

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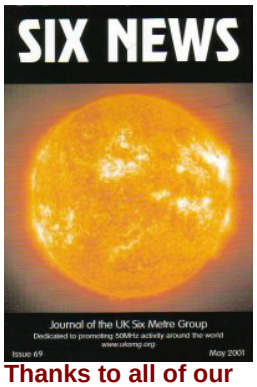
Last Revised: September 2000

The UKSMG Six News Archives



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A link to another 6 meter HELIAX DUPLEXER http://www.qsl.net/kf6yb/duplexer.html

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Amateur Radio

Amateur Radio

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  - Copper Cactus (Multi-Band J-Pole) With Drawings
  - Single Coax Feed (for Copper Cactus)
  - The Stacked-J (Double High J-Pole) With Sketch
  - Mirrored-J (Triple Height, Side Mounted J-Pole)
  - K-FACTOR Chart & Table (Element Length Adjustment)
- Miscellaneous Data
  - National 10-Codes
  - Just Plain NonSense

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Computer Terms

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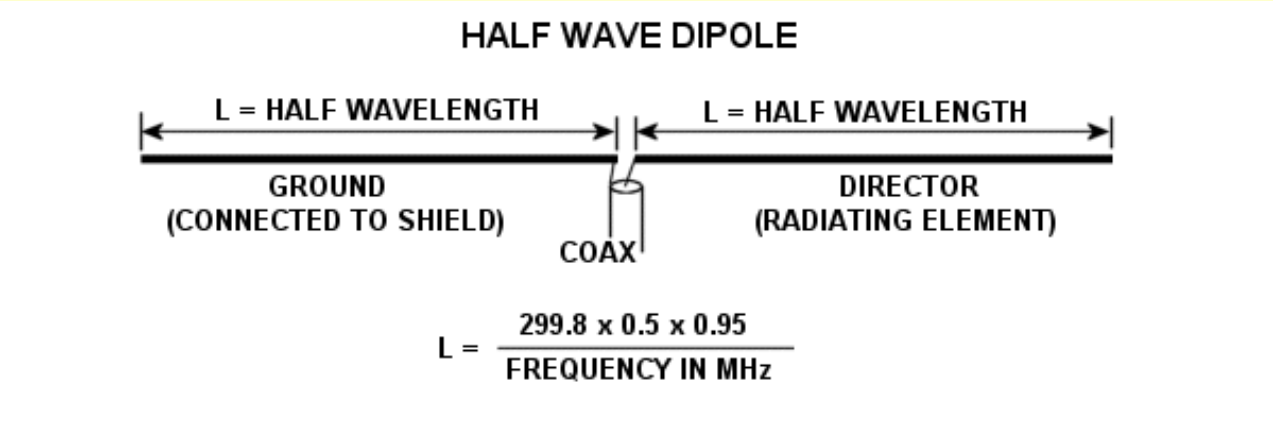
SIMPLE MADE DIPOLE FOR 6 METERS REVIEW BY JIM BAUDO



From: "Rune Austefjord" <rune.austefjord@hjemme.no>

To: "JIM BAUDO" <n0uqz@yahoo.com>

Subject: Re: Homemade 6m dipole

Date: Fri, 24 Jan 2003 17:40:22 +0100

My Quck Review Of My 6 Meter Dipole :

Made a simple dipole from 1.25mm diameter copper speaker wire.
Each of the two elements are (299.8x0.5x0.95)/frequency=142.5cm (for 51MHz).
These elements were soldered to the center and ground of a BNC female connector in this instance.
There were no baluns in use.
SWR was measured to approx. 1.1:1 max in the band 50MHz to 52MHz.
The antenna was hanged vertically in free air at least 2 feet away from any horizontal obstacles, and 4 feet above ground.
Transmitter power was varied from as low as 5W to as much as 100W, with little variation in the SWR.
Because there is little to none 6m activity in my area I didn't have the opportunity to test against other stations.
Overall impression of the antenna is good, easy to make, easy to make play, mobile as it could be coiled up and put in a pocket for transport, cheap and simple. Check out the graphic.

Hope this is what you wanted. Have made a 2m GP-antenna from an old umbrella as well.

73 de LC3LCT

ANTENNAS TRIPODS TOWERS ROTORS AND ACCESSORIES MAIN MENU

MOST ALL PHOTOS WITH HP CAMERA, BY JIM BAUDO. EMAIL YOUR COMMENTS, OR YOUR REVIEW TO JIMBAUDO@YAHOO.COM HAM RADIO CALL N0UQZ

THE 6 METER OMNI HALO

from
KB1DIG

April 28th, 2002, By Steve KB1DIG

Something a little more advanced. Always fun to try out new bands. This halo is made with a true Gamma Section this time and is fashioned from aluminum. Most of the parts are leftovers from old car projects. The 3/8" fuel line I used came from Summit Racing Equipment: #SUM-G2538, and a 25' section costs only about $20.00.

Some of the mods I came up with were:

Welded the elements to the aluminum plate with some of that "Alumalloy" stuff advertised on television. I drilled a small hole in one of the elements to allow condensation to evaporate.

Capped off the end of the gamma arm with a plug to keep the weather out.

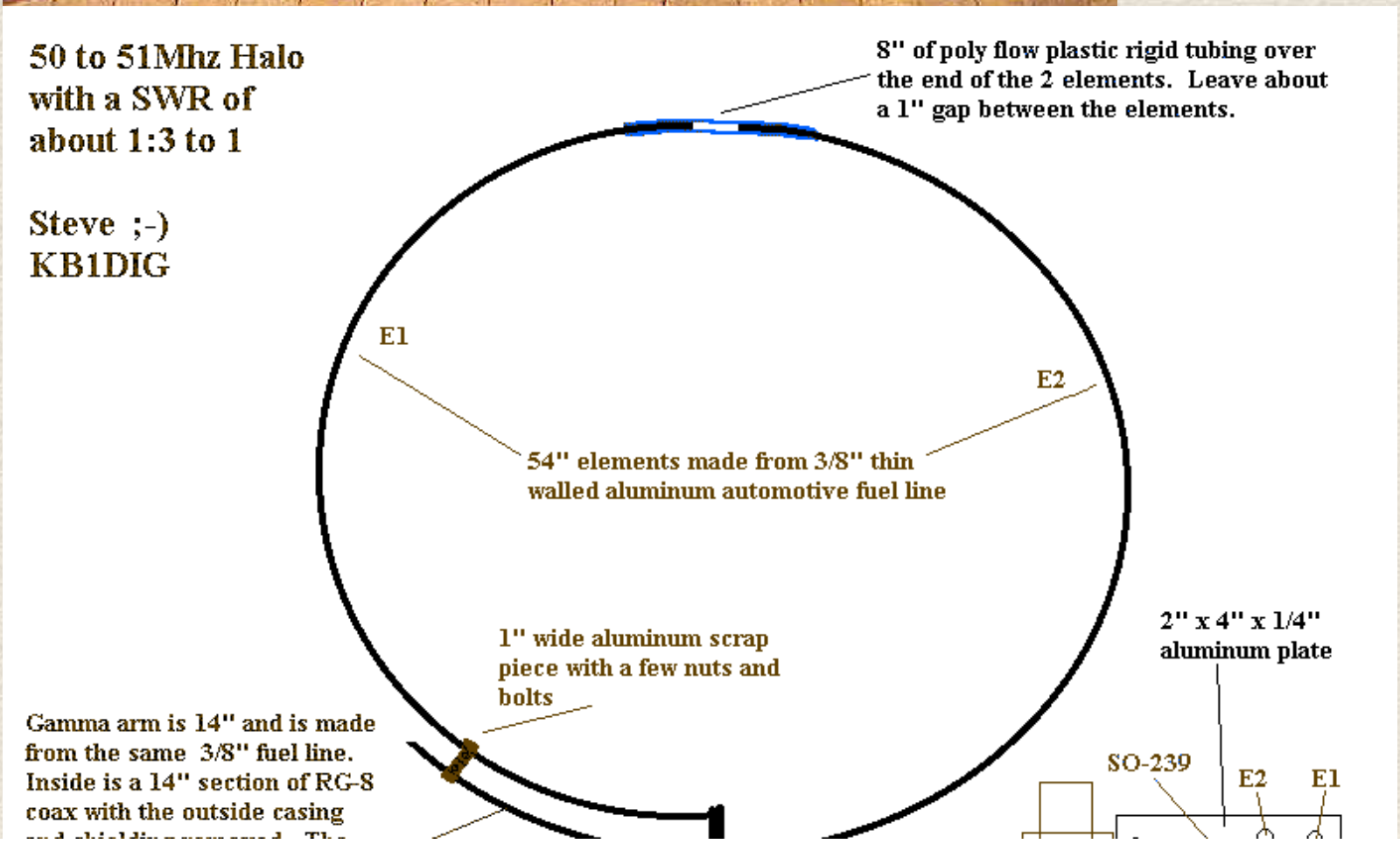
The plug was an automotive type used to block off a PCV line from a carburetor.

After mounting horizontally to a 10' mast I added a support system made from 2 thin 3' fiberglass rods and some wire-ties.

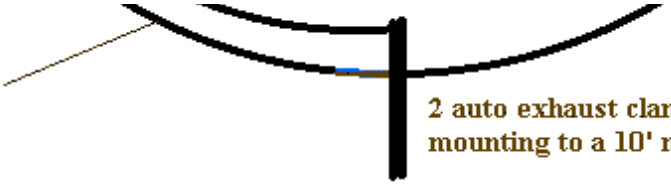
Also remember to hot-glue the wire-ties to the fiberglass rod and adjust for lowest swr as usual.

Works slick! CQ CQ DX!!!

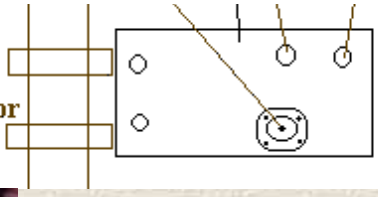
HAM UNIVERSE

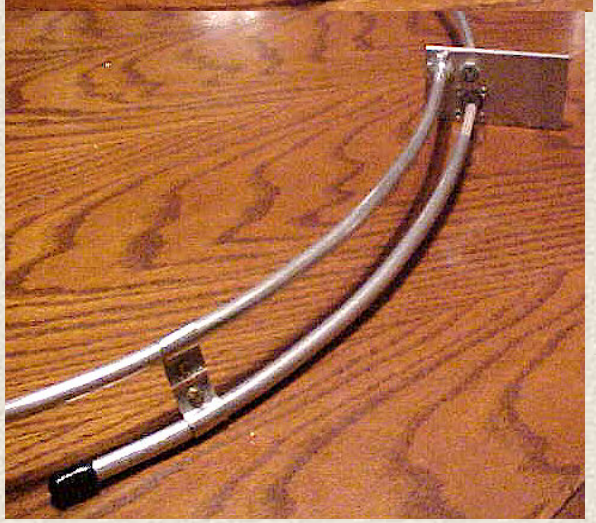


coax with the outside casing and shielding removed. The gap between the Gamma and E1 is about 1 1/2" or less.



2 auto exhaust clamps for mounting to a 10' mast





See closeup #1 below

See closeup #2 below

<<See closeup #3 below>>

Closeup #1

Closeup #2

<<<SO 239

Closeup # 3

The Finished Halo ready to serve!
Pass the rig and a little salt and pepper please!
Anybody want fries!...KB1DIG Have Fun!

<<Notice 1" gap

<<Shorting strap

RG-8>>>

Many thanks to Steve, KB1DIG for allowing us to share his project with you.

Check out Steves site and get questions answered at the link below

"KB1DIG HAM SITE"

Congratulations to Kim, Steve's XYL, KB1GTR, ON HER TICKET!

GOOD LUCK AND 73

N4UJW

BACK TO ANTENNA DESIGN LAB!

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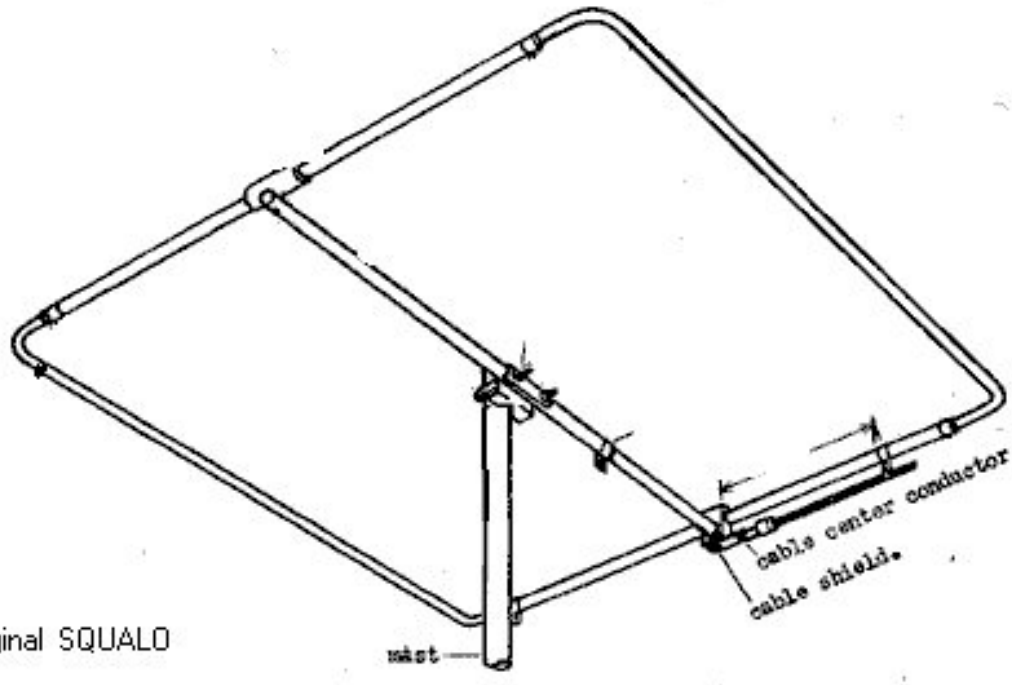
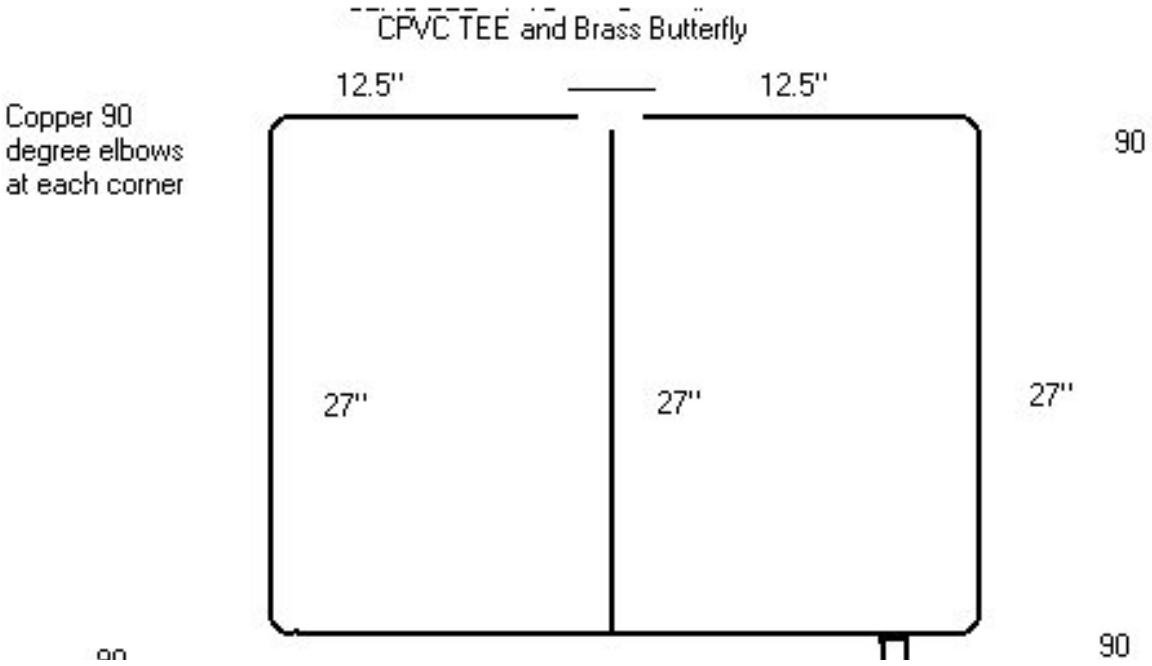
Antennas & More for HF antennas

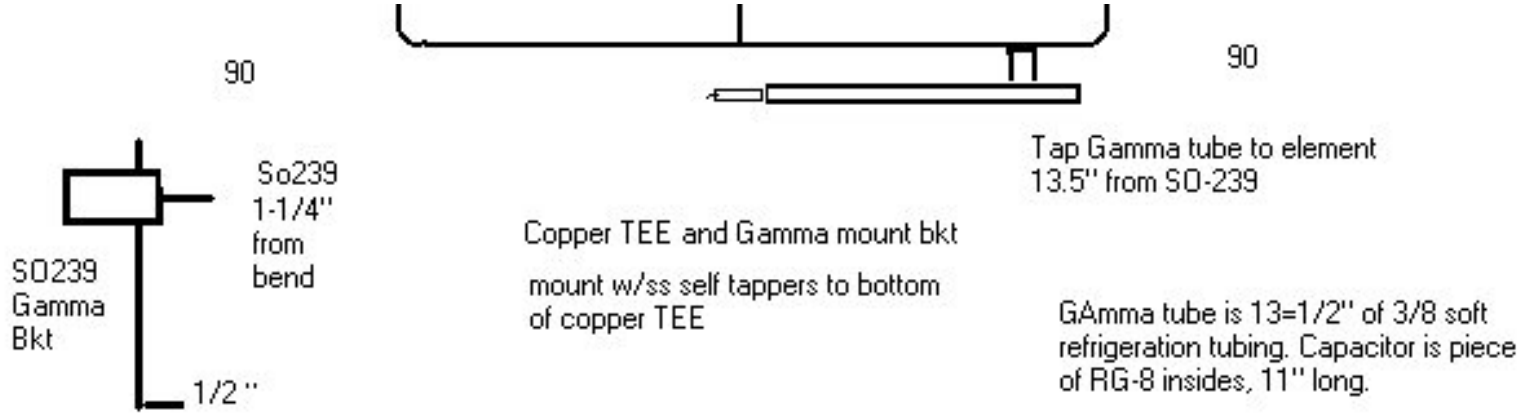
K0FF's 6 -meter SLOOP



Original SQUALO

Electrical Equivelent Circuit
of K0FF Copper Square for 6





Characteristics:

It is more or less omni-directional, and horizontally polarized. Copper is the best possible electrical conductor at normal temperatures, next to silver. Copper conducts better than gold! Antenna efficiency is the RADIATION RESISTANCE of the antenna, divided by the ELECTRICAL RESISTANCE. An antenna made from copper is 1.6 times more efficient than the same antenna made of aluminum.
It presents a high angle of takeoff when mounted low, and singly (great for Es), can handle 100W. Stack two or more for extra gain if needed.

It's just a dipole folded around on itself, and supported at the far (open) end with a plastic insulator. "Bent Dipole" might be a good term, as a "Folded Dipole" is quite antoher thing.

The Gamma Match:

A low VSWR may be obtained by adjusting the Gamma Match shorting bracket position, and also the length of the tubing and shorting bracket. The Gamma bracket "finds" the 50 Ohm point along the element, and connects that to the Coax connector via the Gamma tube. The additional length of tube adds inductance into the circuit, and this is canceled out by the series capacitance formed between the insulated Gamma wire and the inside of the Gamma tube.

Description:

The shape is a closed 28" square, with a mounting /support bar through the middle. This mounting bar is attached to a copper TEE at the drive end, and to a CPVC TEE at the other end. The CPVC TEE acts as support and end-insulator for the radiating element and provides a mounting point for the Butterfly.

A large (3/4") part is used, and adapted down to fit the water pipe, to increase its insulating qualities, as there is very high voltage at this point. A strip of brass or copper 1/2 by 3" is screwed

to the outside middle portion of the CPVC Tee, through a small center hole, and is rotated one way or the other as a resonance tuner (Butterfly). When the Butterfly is at right angles to the element, the frequency is the highest, when parallel, it's the lowest.

A Gamma match sets the impedance to 50 Ohms, and the Butterfly adjusts the center frequency. Center frequency is 50.00 to 50.800 with the exact dimensions shown. Typically the 2:1 SWR bandwidth exceeds 500 kHz.

Mounting:

A U bolt and saddle through the central tube provides a center mounting point. Another approach is to install a copper TEE in the center tube, with the open end down. In that open end solder a 1/2" brass rod which has been drilled and tapped for 3/8-24.

Side mounting on a tower can be achieved by using conduit clips to fix it to a horizontal mast.

In some climates where water is a problem, drill small weep holes in the bottom corners. A spray coat of Krylon Clear Enamel will keep the copper shiny. If used mobile, you may use a colored paint, the same shade as your vehicle.

Construction:

Material:

1/2" Copper waterpipe :

3 ea. 27 inch piece
4 ea. 12.5 inch piece
4 ea. Copper 90 Degree elbow
1 ea. Copper TEE

Other Copper or Brass :

1 ea. 13.5 inch piece 3/8 i.d. Copper refrigeration tubing (Gamma tube) *
1 ea. Brass plate 1/2" x 3" (Butterfly) *
1 ea. Brass plate: 1"x 2-1/4 in (to mount SO-239, Gamma rod) *
1 ea. Copper strip 3/8 x 4" to make Gamma tube bracket *

PVC :

1 ea. 3/4 CPVC TEE

3 ea. 3/4 to 1/2 CPVC reducer

Odds and Ends :

1 ea. 11" piece of RG8 insides (center conductor and insulation, Discard shield and outer plastic) *
1 ea. SO-239 coax connector (with tapped mounting holes and center pin) *
1 Lot Stainless Steel and Brass screws and Hardware *

Glue two of the 12.5" pipe sections into the CPVC reducers first, then glue the reducers into the opposite sides of the CPVC TEE. Lay the assembly on a flat surface with the center opening of the TEE facing the middle of the antenna. This is where the first 27" piece (mounting bar) goes in, via a reducer. On the outside edge of the CPVC TEE is where the butterfly attaches. For mobile or portable use, use 3 s.s. #6 screws through each of the CPVC TEE joints for added strength. The rest of the antenna solders together to form a square, using the 90 degree elbows at the corners. Drill small weep holes in the bottom corners of all four 90s to let accumulated water drain out.

The brass plate is bent to form an "L" 1-3/4" tall with a 1/2" lip. A 5/8" hole is provided 1-1/4 inch from the bend, and an SO-239 is attached . Two small holes are drilled in the lip and the plate is mounted to the copper TEE with s.s. #6 self taping screws. Attatch the center conductor of a 11" piece of RG8 insides to the center pin (center wire and plastic dielectric only- remove and discard shield and outer covering). This is accomplished by soldering or using a screw if the SO-239 has a threaded center pin *. Slip the other end of the RG8 insides into the 3/8"copper tube 10.5", and tap the copper tube to the radiating element 13-1/2" from the SO-239 center, with the Gamma tube bracket.

The tap on the Gamma sets the impedance presented to the feedline. Resonance (center frequency) is adjusted by turning the butterfly.

Mount 15 feet or more high, for home use, and wherever you can for mobile. An antenna like this can be mounted 3" to 6" above the roof of a vehicle using CPVC , PVC or acrylic spacers with suction cups.

Have fun on 6. Geo>K0FF

***Parts available in a kit from author, includes all "*" parts, drilled, punched, bent, and threaded, contact:** K0FF@ARRL.NET

Don't Eat the Batteries" clause:

Recently my wife and I bought a new TV set, and in the instructions for the hand held remote control the warning said "Don't Eat the Batteries".

SO....

WARNING!!! **This is antenna is an electrical conductor. Contact with power lines can result in death or serious injury. Do not install this antenna, supporting mast or tower structure near any power lines, or where they could come into contact with power lines should the antenna or structure fall.**

Geo

Return to Home Page.

A Long-wire Aerial for Six Metres

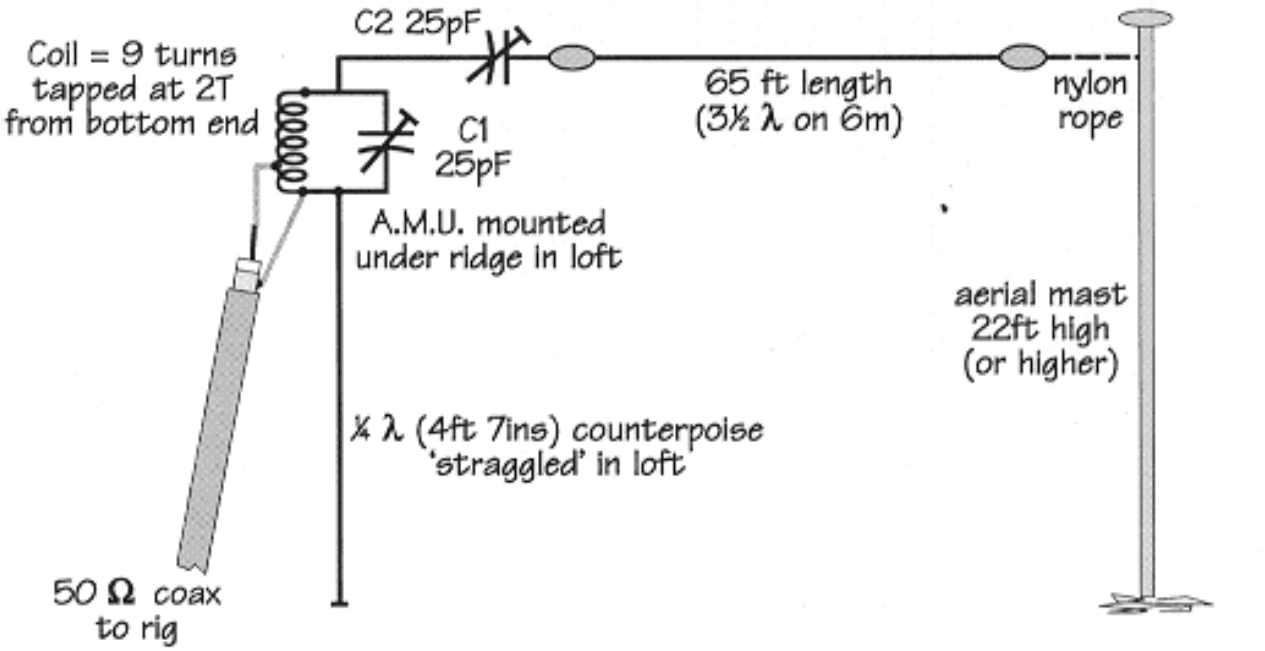
Brian D. Williams, GW0GHF

Issue, 56 February 1998

Having had the time to evaluate it, I find this aerial very useful for both semi-local and sporadic-E on six metres.

The antenna is about 65 feet of ordinary aerial wire, insulated at the far end. The near end connects to a matching unit situated in the loft of the house. The aerial wire enters the loft under a tile near the ridge of the roof of the house, through an insulating sleeve.

The



theoretical diagram.

The aerial is voltage-fed (high impedance) and the antenna matching unit (AMU) transforms this to match 50 ohm cable. The two trimmers are 'polycon' variables - these are OK up to 50 watts. For operation at higher powers, obviously more substantial trimmers, such as mica compression types, will be needed. Old radio receivers of the 1950 vintage or before are often a source of these types of trimmer. 'Cirkit' can supply them, but they are not cheap! The trimmers are likely to be the only 'expensive' parts required for this antenna, however.

The method of adjustment is to set C2 to half capacity and then adjust C1 for best noise on receive. Then feed in a carrier (or an audio tone on SSB) and adjust C1 for a low SWR reading. 1:1 is achievable at this QTH.

No problems with 'RF in the shack' have been encountered at 50 watts SSB. The counterpoise were effectively 'earths' the bottom end of the parallel-tuned circuit; it hangs from the AMU in the loft. It is 4ft 6in in length.

The AMU is constructed on a tobacco-tin lid. The coil is nine turns of 18 SWG copper wire taken from old coax cable. It is 3/8 inch in diameter, spaced over approximately ¾ inch.

Increasing C2 too much will prevent the tuned circuit from resonating at 50MHz. This, in turn, depends on the capacitance of the whole aerial wire to earth - but no problems were found and it was easy to get unity SWR. Shorting out C2 might be acceptable if a low-capacity system can be constructed, but this did not work at my QTH.

The wire gives a gain of 4 dB in four directions at 35 degrees to the run of the wire. There are also several smaller lobes which are useful. This is a good ‘listening’ antenna with gain that can be arranged in desired directions.

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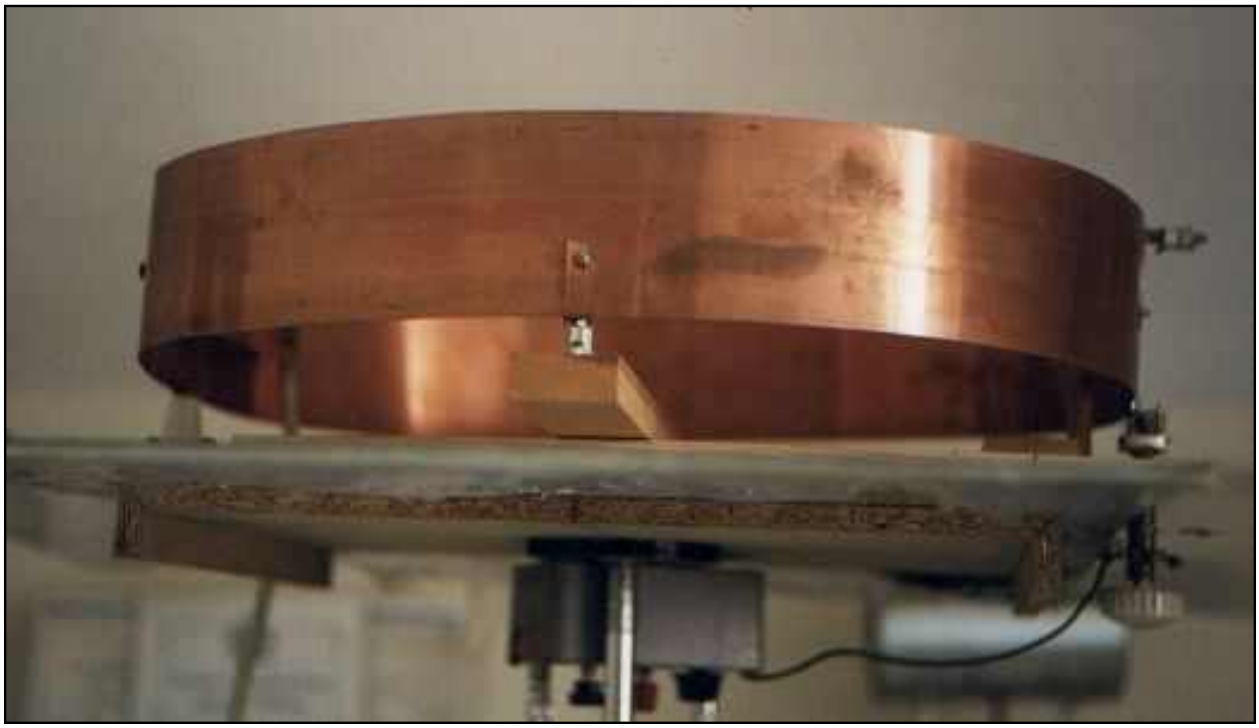
An Indoor Loop Antenna for 6m

Colen Harlow, G8BTK

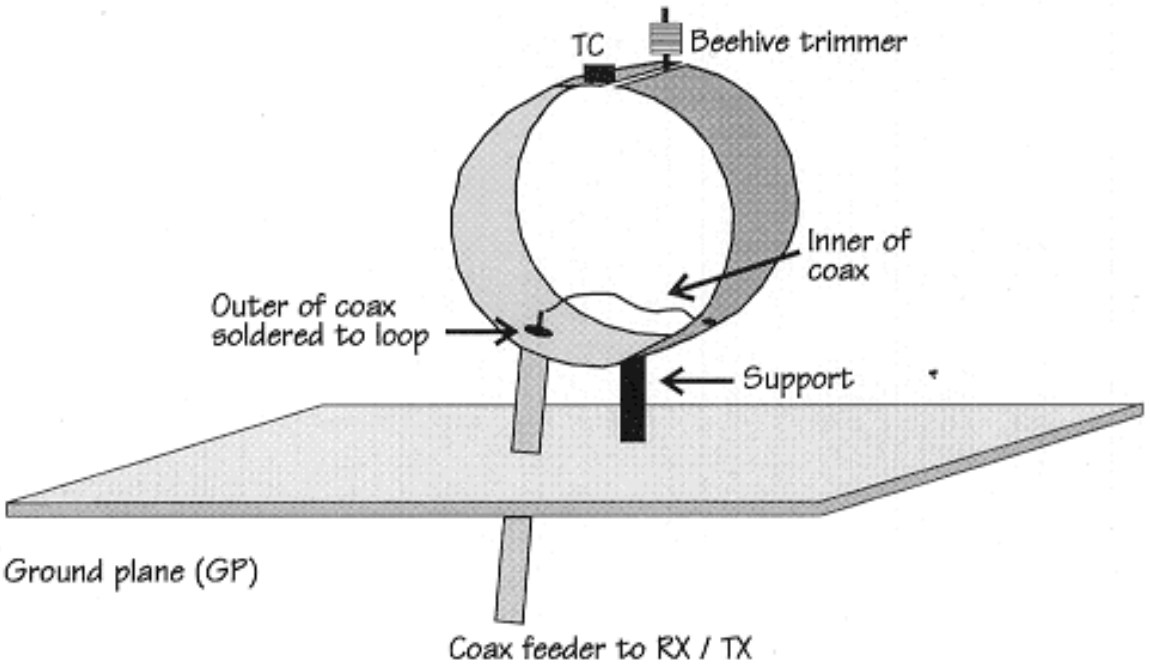
Issue, 56 February 1998



The 6m 20" high Loop.

I have worked 24 countries with my 20 inch loop. Living as I do in a flat with a complete ban on external antennas, I turned to the loop, having for many years experimented with this type of aerial in various forms. I now use them for six metres (20" diameter), four metres (10" diameter), two metres (6" diameter) and 70 centimetres (3" diameter).

G8BTK's

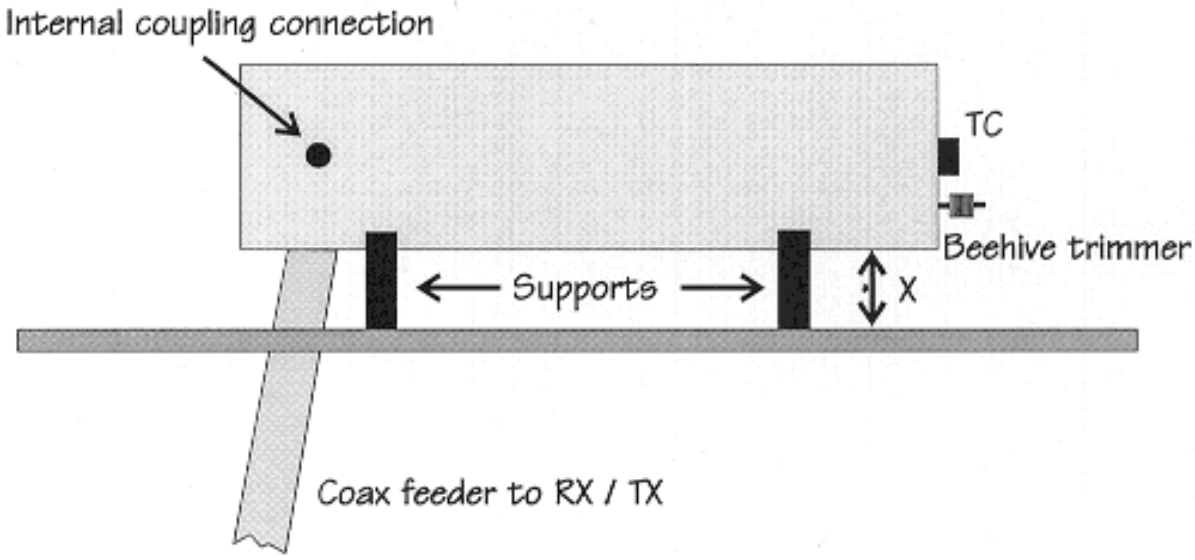


vertically polarised loop

The six metre loop (figures 2 and 3) consists of a four inch wide strip of copper formed into a ring, the ends of which are joined by a butterfly trimmer and beehive trimmer in parallel, for ease of adjustment of resonance. The loop is mounted above a ground plane of a three foot square aluminium sheet - hardboard covered with kitchen foil would do just as well. The loop is centrally positioned, about four inches from the ground plane.

The feed-point is taken at right angles from the opposite side to the capacitors. Good quality (ceramic, PTFE or similar) insulators should be used. The feeder coupling consists of a strip of copper or heavy-gauge wire as shown in the diagram.

The



horizontally polarised version of G8BTK's loop

The separation of the loop from the ground plane ('x' in the diagram of the horizontally polarised version) determines, to a great extent, the radiation angle, as does the size of the ground plane. With a three foot ground plane I have found between four and 12 inches to be the best.

6m Loop Coupling and feed

For six metres a resonated loop also works very well and the vertically polarised version exhibits a very pronounced null.

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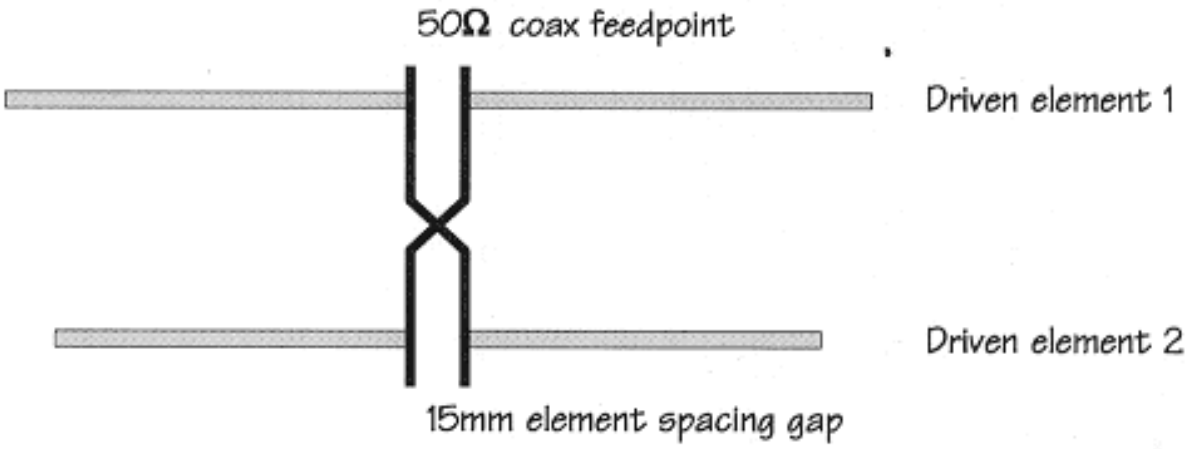
The ZR6YY Antenna for 6m

Maurius, ZR6YY

Issue, 56 February 1998

The



Driven elements of the ZR6YY antenna

Having received numerous requests to publish the design for the ZR6YY yagi I have now put the information out world wide in the hope that the design works for you.

Element Lengths	
reflector	320 cm
driven 1	296.5 cm
driven 2	274.5 cm
director 1	255 cm
director 2	254 cm
director 3	245 cm

Element Spacing	
The measurements are taken from the reflector element	
reflector	0
driven 1	78.5 cm
driven 2	120.8 cm
director 1	179.2 cm

director 2	267 cm
director 3	383 cm

Construction

All the elements are isolated from the boom. The two driven elements are fed direct and 180 degrees out of phase. Feed the antenna directly with 50 ohm coax no balun necessary. Driven element 1 and driven element 2 are made up of two pieces of aluminium each separated by a 15mm gap.

Technical note from Clive, G4FVP

The design shown has considerable experimental potential and should give a good forward gain on a short boom length of only 3.83 metres. The concept of the yagi is the two driven elements which act as an end-fire array. This should theoretically give a gain of 3dB but will in practice be less. Remember to reverse the feedline between the two driven elements to ensure they are 180 degrees out of phase.

Maurius does not specify the feeder between the two driven elements however a starting point would be to use an electrical half wave length of 300 ohm slotted ribbon feeder. To determine the electrical half wave multiply the free space length by the velocity factor of the cable in use. For 300 ohm slotted ribbon feeder this will be 150/f X 0.87. For 50.100MHz this is 2.60 metres.

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Moxon Rectangles for 6 Meters

L. B. Cebik, W4RNL

I have had numerous requests over the last few years for the dimensions and construction plans for Moxon rectangles designed for the 6-meter band. The Moxon rectangle is a quite broad-band antenna, but it is not quite broad enough to cover the entire band. As well, the lower end of the band is the major arena for horizontal polarization using CW or SSB. The upper portion of the band sees most of the FM activity, with vertical polarization being standard.

Hence, for full band coverage--or for selected use of one or the other mode of activity--we really need 2 Moxons. The first will be a horizontally oriented beam designed for 50.5 MHz with coverage of the first MHz of the band. The second will be a vertically oriented version designed for 53 MHz, with coverage from 52 to 54 MHz. After looking at the characteristics of these two versions of the same basic design, we shall make a few construction suggestions. Finally, we shall show how to combine them into a single array--but with separate feedlines.

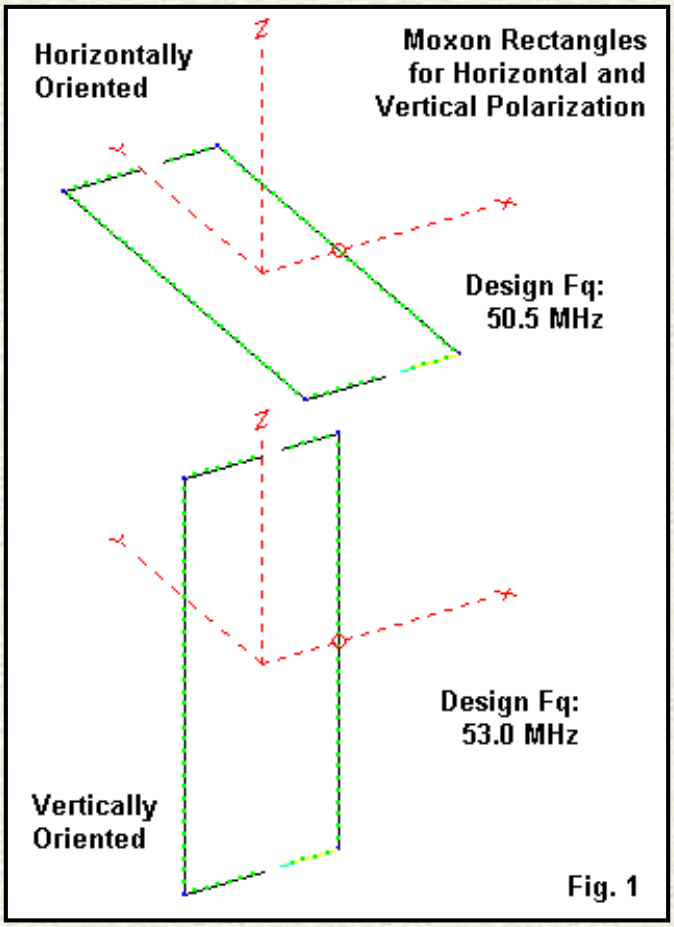
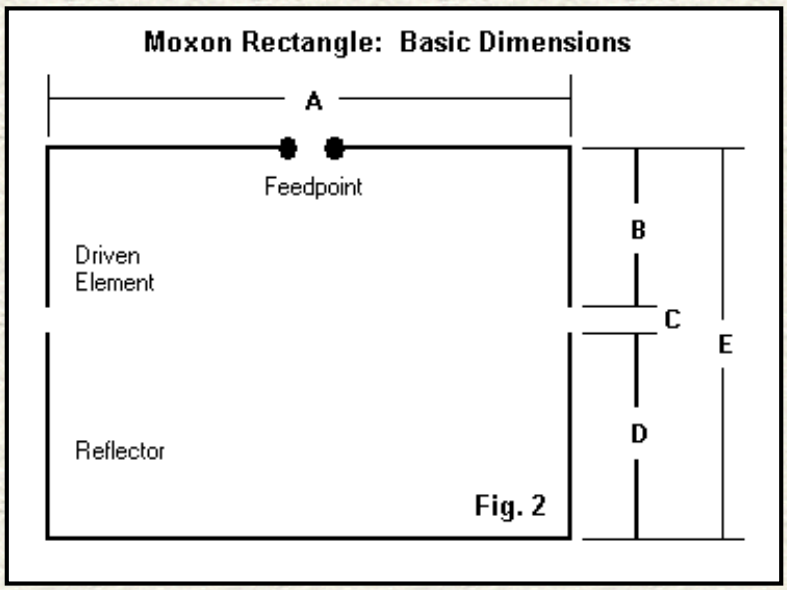


Fig. 1 shows the general outlines of the two types of Moxons. The Moxon is a driver-reflector type of parasitic array. Unlike standard Yagi designs that employ only the coupling between parallel lengths of element conductor, the Moxon folds back its elements to provide a second form of coupling. The coupling of the driver and reflector tails that face each other provides a second form of coupling, and the combination of the two gives us an array that we can design for very good front-to-back performance, about as much forward gain as a 2-element Yagi, and a 50-Ohm feedpoint impedance for direct connection of

a standard coaxial cable feedline.



Since we need a way to refer to the parts of a Moxon when giving dimensions, **Fig. 2** supplies what has become a set of default designations. Dimension A is the total side-to-side dimension of the array. B is the driver tail or fold-back portion, and D is the reflector fold-forward portion or tail. C is the most critical dimension, the gap, and tends to vary as a direct function of the element diameter. E is simply the sum of B, C, and D, and gives us the total front-to-back dimension of the array.

All 6-meter Moxons will be about 7' wide or about 3.5' each side of the center line. The front-to-back dimension with be about 2.5', plus or minus a little. Hence, the Moxon makes a very compact array, suitable for enhancing repeater communications or for SSB operation in local nets.

In fact, the precise dimensions for a 50-Ohm Moxon for any frequency and element diameter have been developed into several computer programs, ranging from a GW Basic utility in the HAMCALC suite to a NEC-Win Plus model to a stand-alone Windows program developed by AC6LA and available for free download from his site (**http://www.qsl.net/ac6la**). Since all of them are based on the same modeling and regression analysis that I performed some time back, all will give the same dimensions for the same design frequency and element diameter.

Let's start our foray into 6-meter Moxons with the low-end horizontal version.

A Horizontal Moxon Rectangle for 50.5 MHz.

The materials that folks have access to will vary from region to region. Therefore, let's make a chart of dimensions. All of the dimensions will presume that we are using some form of aluminum tubing, ranging from 1.0" down to 0.25" in diameter. As we shall see in the construction section, aluminum tubing is an optimal choice for a 6-meter Moxon.

In the following table, all dimensions refer to **Fig. 2** and are in inches.

. . . . . .

Dimensions for a 50.5-MHz Moxon Rectangle

El. Dia.	A	B	C	D	E
1.0	83.61	10.40	4.58	16.22	31.20
0.875	83.68	10.53	4.46	16.21	31.20
0.75	83.76	10.69	4.32	16.19	31.20
0.625	83.86	10.86	4.16	16.18	31.20

0.5	83.97	11.07	3.97	16.15	31.19
0.375	84.12	11.32	3.74	16.12	31.19
0.25	84.31	11.65	3.44	16.08	31.18

. . . . . .

Note that the dimensions change only a small amount from one tube diameter to the next. Moreover, the front-to-back dimension (E) changes almost not at all. However, the differences are important to centering the performance curve of the Moxon on the design frequency, which then has consequences for performance at the band edges. So using the dimensions that apply to the element diameter that you will use does have significance.

Let's set the antenna 25' above ground, which is just over 1.34 wavelengths up. The forward gain will be about 11.4 dBi at 50.5 MHz, with a 180-degree front-to-back ratio of about 30 dB and a 78-degree beamwidth between -3-dB points. The feedpoint impedance of 50 Ohms, plus or minus 1 to 2 Ohms reactance.

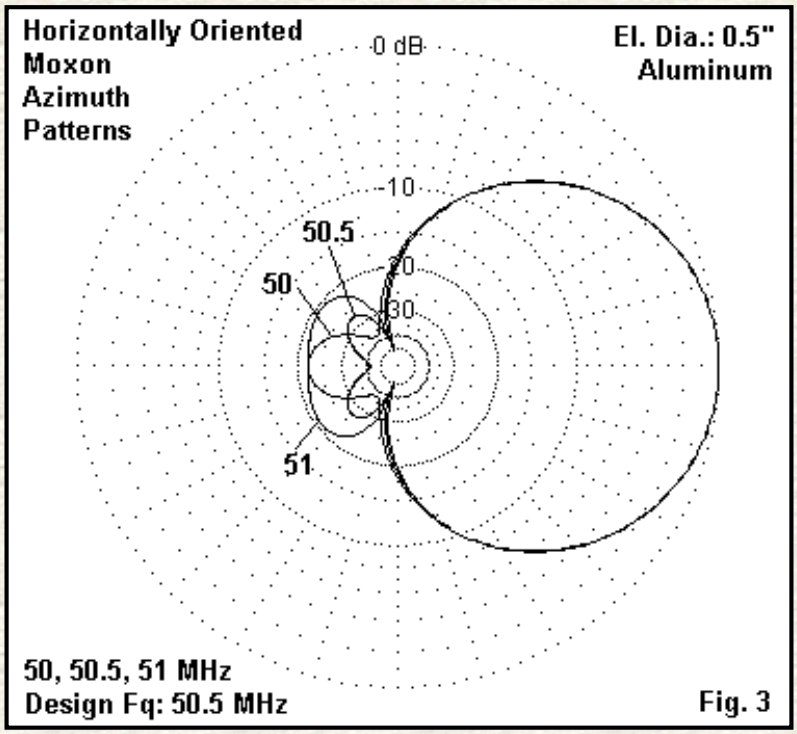


Fig. 3 overlays the design frequency and the passband edge azimuth patterns of the Moxon rectangle when oriented horizontally. As you can see from the patterns, taken for a version using a 0.5" diameter element set, as we move lower in frequency, the gain increases very slightly (too slightly to ever be measured in operation), and the rearward radiation begins to increase. Above the design frequency, the gain decreases by an equally slight amount, and, again, the rearward radiation pattern shows growth.

The performance of the antenna is virtually unchanged at the design frequency for any tubing size. However, the band-edge performance does change (in this case, using a 1-MHz passband). The fatter the tubing, the slower the rate of forward gain change. More significantly, the fatter the tubing, the slower the growth of rearward radiation lobes both above and below the design frequency. However, those changes are not so great as to override considerations such as the most convenient tubing size for constructing a Moxon rectangle.

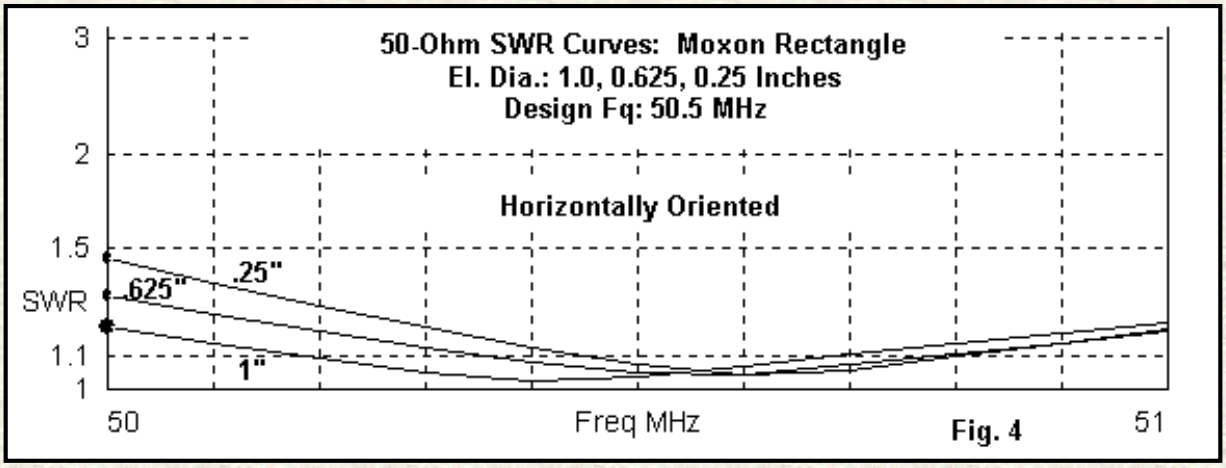


Fig. 4

Tubing size also makes a difference in the 50-Ohm SWR for the final antenna, as measured at the antenna terminals, as shown in **Fig. 4**. For a 1-MHz passband, almost any size tubing will do, and the SWR at the shack end of the coax is likely to be too low to get a definite frequency for the lowest value.

A Vertical Moxon Rectangle for 53.0 MHz.

The design principles do not change at all when we flip the Moxon rectangle for upper 6-meter service. However, the dimensions will change, since we are now using a design frequency of 53.0 MHz in order to cover the 52-54-MHz range. The following table provides dimensions, again in inches and again using **Fig. 2** as a reference.

. . . . . .

Dimensions for a 53.0-MHz Moxon Rectangle

El. Dia.	A	B	C	D	E
1.0	79.64	9.86	4.41	15.46	29.73
0.875	79.71	9.99	4.29	15.45	29.73
0.75	79.79	10.14	4.16	15.44	29.73
0.625	79.88	10.31	4.00	15.42	29.73
0.5	79.99	10.50	3.82	15.40	29.72
0.375	80.13	10.75	3.60	15.37	29.72
0.25	80.31	11.07	3.31	15.33	29.71

. . . . . .

The last decimal place in column E, the overall front-to-back dimension, may be a digit or two off the sum of B, C, and D due to rounding of the individual values. However, I doubt that any builder will be constructing the elements to a hundredth of an inch tolerances. In fact, in the construction section, we shall be slightly altering the dimensions to take account of the fact that we shall bend the tubing at the corners.

Once more, let's place a 0.5" diameter version of the antenna at a height of 25' above ground. At the design frequency, we shall obtain a 50-Ohm feedpoint impedance accompanied by a front-to-back ratio well above 30 dB. (Vertical orientation affects the front-to-back ratio less than horizontal orientation for an antenna within about 2 wavelengths of ground.) However, unlike the horizontal version of the array, the peak gain is only about 7.6 dBi. Let's see why.

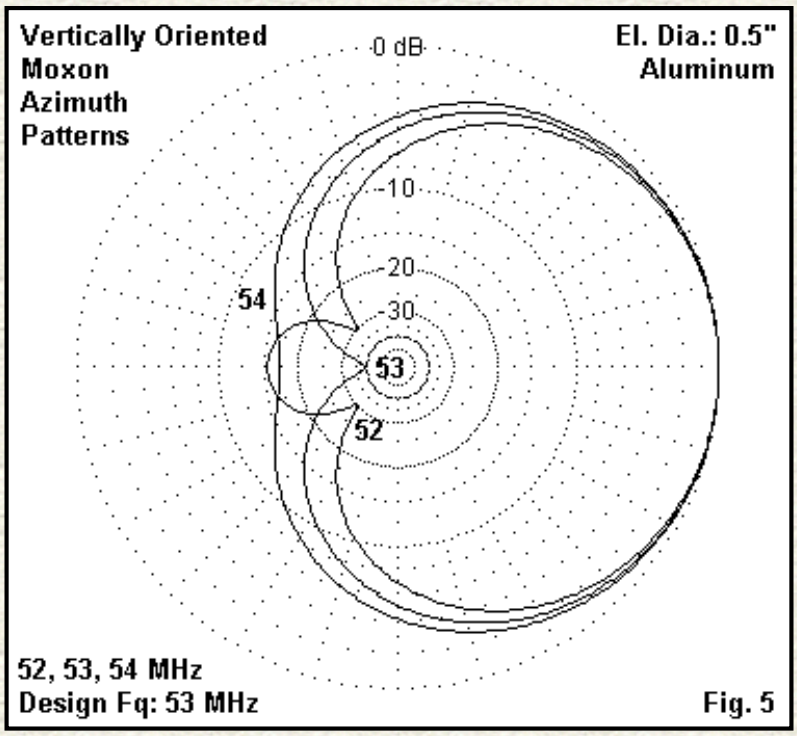
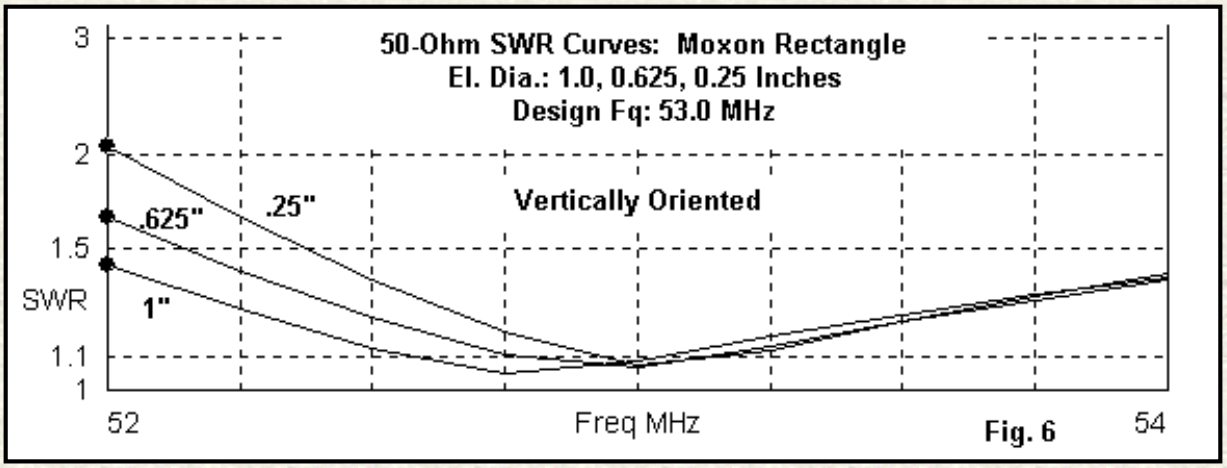


Fig. 5 overlays the azimuth patterns of the antenna at the design frequency and at 52 and 54 MHz. In all cases, we see a very wide beamwidth, over 142 degrees between -3-dB points. That increased beamwidth--about twice the value for the horizontal version--spreads the radiated power over a much wider area and thus reduces the peak gain. This is inherent in any parasitic array with all of the elements in a single plane. Nevertheless, both the horizontal and the vertical versions of the array have almost 4 dB gain over their counterpart dipoles at the same height.

As the patterns show, the increased bandwidth that we require of a vertical Moxon increases the rearward radiation at the band edges. Once more, the fatter the elements, the less growth to the rearward radiation for any change in frequency relative to the design frequency.



In **Fig. 6**, we have the 50-Ohm SWR curves for the array at selected element diameters. The SWR increases more rapidly below the design frequency than above it. The antenna feedpoint impedance for the thinnest tubing actually is above 2:1, although at the shack end of the coax, where we usually measure SWR, it may seem lower than that value. Hence, one might well think in terms of at least an intermediate tubing size for the vertical Moxon.

The SWR climbs more slowly above the design frequency, suggesting that we might choose a lower design frequency and extend the coverage from 51 to 54 MHz. This tactic is possible, but at a cost. The forward gain and the front-to-back ratio degrade continuously as we move above the design frequency. Hence, with a design frequency of, say, 52 MHz, the array

performance would not be very good at 54 MHz.

Building a Moxon Rectangle for 6 Meters

There is an unfortunate tendency among newer antenna builders to see a design they like and then to grab almost any materials close at hand and slap together a version that almost works. Many a good antenna design has gotten a bad name in some regions because builders did not exercise the same care in construction as the original builder. To obtain performance that agrees with the design notes above, acquire the right materials and then build the antenna with all the care possible.

A good tubing size for a 6-meter Moxon--whether horizontal or vertical--is 0.5". This size is useful, since we can use #8 or #10 hardware for fastenings. Of course, the hardware will all be stainless steel, both for rust prevention and to avoid bi-metallic contact problems.

For a 6-meter Moxon rectangle, we shall need 4 6' lengths of 1/2" diameter tubing. This size tubing can be shipped from suppliers like Texas Towers by UPS. (Yes, we shall a some scrap left over for use as garden stakes.) Use 6061-T6 or 6063-T832. Hardware depot tubing has an unknown vintage, so good quality antenna tubing is highly desirable. Do not use aluminum electrical conduit or copper tubing. The conduit is too heavy, and so is the copper in any form rigid enough not to gradually fold over on its own accord.

We shall also require a short (under 6") length of 3/8" aluminum tubing and a similar length of 3/8" diameter fiberglass or similar rod. We shall be constructing the elements in halves, so we need to join and align them. The short length of 3/8" aluminum tubing will join the two halves of the reflector, making them electrically one. The fiberglass or equivalent rod will align the driver halves, but allow a gap for connecting the feedline. Finally, we shall require some 3/8" outer-diameter fairly rigid tubing, something light but straight. These tubes will fit inside the ends of the driver and reflector tails to hold the spacing constant under all conditions.

For hardware, we shall require some #8 nuts and bolts, along with some locking washers. We can use sheet metal screws to fasten the tail junction tubes in place. However, all hardware must be stainless steel, including the washers. Since some of this falls outside what home warehouse hardware bins contain, consider locating a hardware supplier or use an on-line ordering source like McMasters-Carr. From such sources, you can also obtain a small sheet of 1/4" thick UV-protected polycarbonate (trade-name Lexan) to use as boom-to-element plates. Polycarbonate cuts nicely with woodworking saws and drills cleanly with standard bits.

The needs for a single Moxon rectangle are small, so you may wish to combine orders with others interested in the antenna in order to make up the minimum order requirements for a given supplier. Since we have vowed to be careful, we need not rush to get parts, but can go slowly and get everything we need.

For a boom, you can use either metal or Schedule 40 PVC (if the PVC in your area is adequately UV protected--this varies around the U.S.). In the southeastern US, white PVC gives me about 10 years of service before becoming brittle.

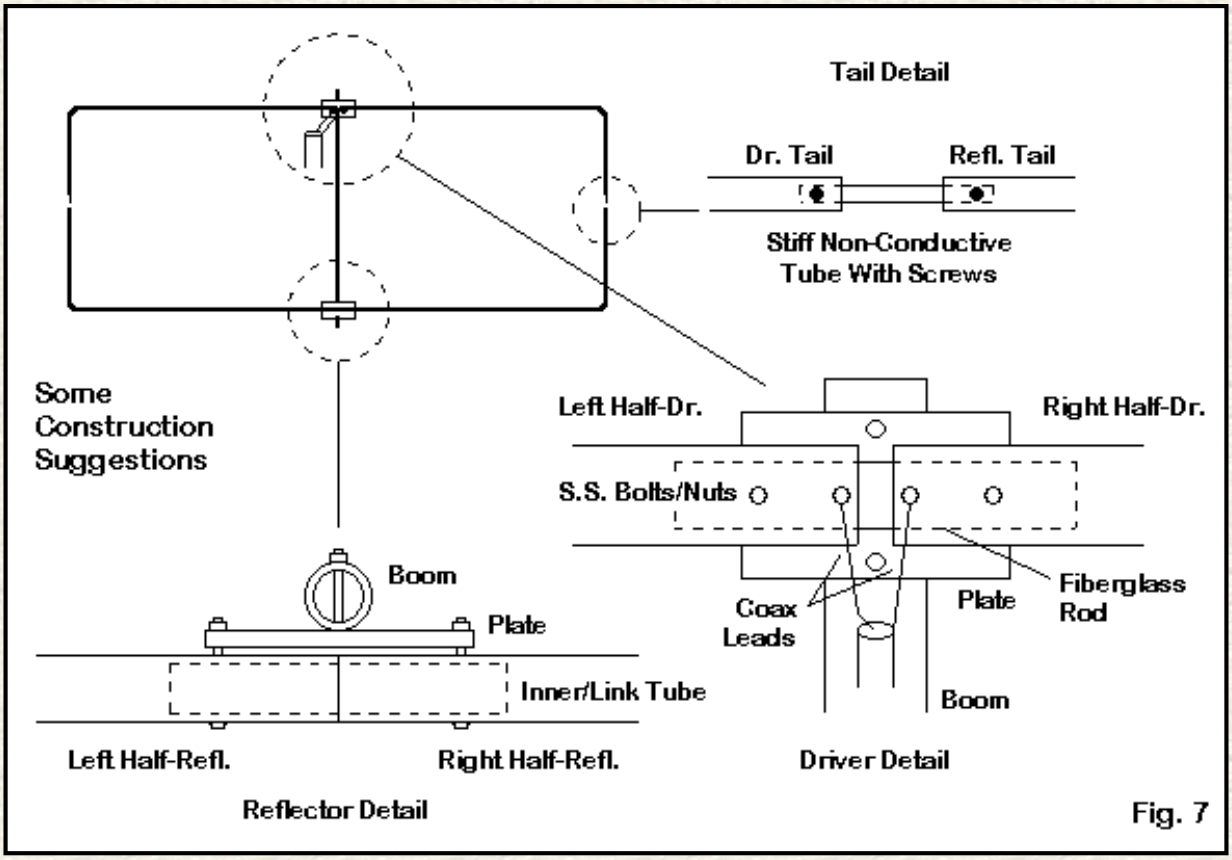


Fig. 7 shows some of the suggested construction methods that I use. you may have better ones, in which case, use them. There are 4 keys areas of construction concern.

1. *The reflector junction*: A 3/8" section of tubing a little longer than the boom-to-element plate joins the two halves. If we use polycarbonate plates about 4" long and 3" wide, we have plenty of room for the elements and the nuts/bolts for boom fastening. A 1" nominal PVC pipe is actually about an inch and a quarter in diameter and is very rigid for a boom that is less than 3' long. For the plate-to-boom bolts, #10 hardware is very secure, using only 2 bolts per plate. However, use a compression lock washer against the plate or obtain self-locking nuts (with a nylon insert).

Since a #10 bolt requires a larger hole, you may wish to fasten the elements to the plate with #8 hardware, again with compression lock washers against the plate. (Toothed lockwashers may gradually loosen by gouging the polycarbonate.) For all drilling of the boom and the elements, make up a jig from scrap wood to pin the material in place while you drill. If you can gain access to a drill press--even a small device designed to hold an electric hand drill--by all means use it. Align all holes before drilling instead of widening holes later to bring parts into alignment.

2. *The driver gap*: By using a 3/8" fiberglass rod to align the driver halves, we can fasten the driver and the rod to the plate using #8 hardware. Note that there are two sets of hardware at the driver: an outer set to pins the element to the plate and an inner set to which we shall connect the feedline. If you prefer, you can set the connection hardware at right angles to the element-to-plate hardware to keep the coax more in line with the boom.

The sketch shows direct connections between the element and the coax, with no connector. I have found that from 6 meters on upward, connectors and their associated leads contribute reactance to the feedpoint impedance. A direct connection and a short length of coax taped to the boom for strain relief simplifies feedpoint construction. Once everything is complete, seal the connections, especially since solder terminals may not be available in stainless steel. Plasti-Dip or similar materials provide a weather-secure coating.

A note on the gap: The gap at the feedpoint is part of the overall element length (or dimension A), NOT an addition to it. Whether you start with a 1/4" gap or a much wider one, let the driver side-to-side dimension remain constant. In effect, the coax leads make up the seemingly missing tubing. The driver gap is not critical from 1/4" to over 3/4", but closer is always

better in this type of antenna.

3. *The tail separation*: Because we must keep the tail ends at a specified distance and aligned, we need a short piece of non-conductive tubing to lock their relative positions. Almost anything will do here, if it is 3/8" in outside diameter and relatively rigid. We shall need only about 4-4.5 inches exposed, so even flexible nylon tubing will work, although rigid plumbing CPVC is superior. You may use sheet metal screws to fasten the tube inside the tail pieces--after careful measurement, of course.

4. *The element bends*: Bending aluminum tubing requires care to prevent crimping that will eventually result in a metal crack and break. The radius of the corner bends will depend on the tubing size used. A plumber's tubing bender is applicable only up to about 1/4" diameter tubing. Larger tubing requires larger bend radii, and that means a home-made jig.

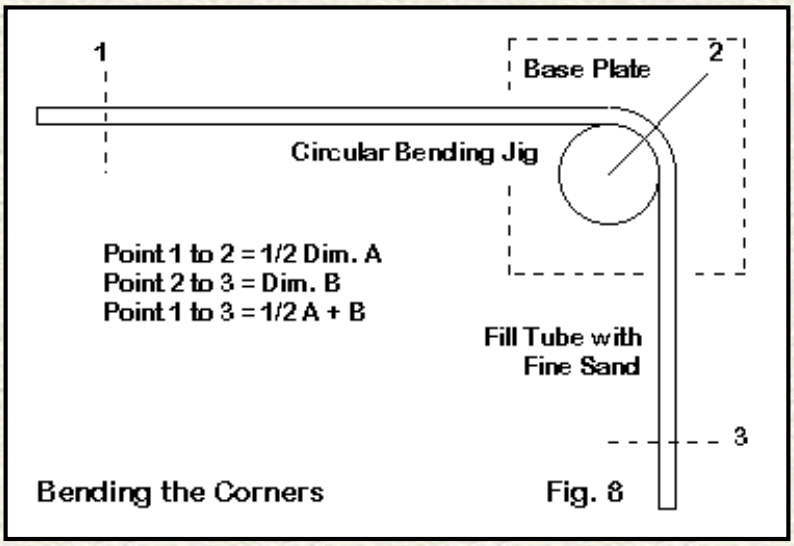


Fig. 8 shows a simple jig: a circle mounted on a base plate. The circle can be cut from plywood or be a pulley wheel. For larger radius bends, I have used such materials as a worn-out power mower wheel.

Mark a point on the 6' length of tubing that marks the center of the bent corner and leaves excess on both the tail and the parallel sections. Try to keep this mark at the center of the bend. As well, mark the tubing to indicate the parallel length (1/2 of dimension A) and the tail (B or D, as applicable). Fill the tube with the finest sand available and tape the end shut. If you prefer, warm the tube until it can just barely be handled with gloves. Pin one end of the tube (a nail in the base board will do) and slowly bend the tubing. The fatter the tubing, the more important it is to bend a few degrees and pause. Continue the bend until it is at least a 90-degree bend. A little more will not hurt, since a slight unbending of the tube does no harm.

Bend the tubing before you drill any mounting holes to make sure that all such holes are at right angles to the mounting plates. In fact, let's delay any drilling and do some work on the floor.

From the dimensions that apply to your tubing size, draw out the Moxon on the floor (or on paper taped to the floor). Be sure to mark the center or boom line as well as the points where the tails end. Next, lay the untrimmed bent pieces on the drawing. Because the corners are "clipped" by the bending, the pieces will just exceed the lines on the floor.

NOTE: For the adjustment of positions, be sure that the gap is constant and does not change!

With a constant gap distance as specified for the tubing size and measured against the marks made on the tubing, adjust the tube position to equalize the amount by which the tubes fall outside the original lines. At the same time, align the tubes. Now trim the tubes. Remember to trim a bit (1/8" or so per tube) off the driver to leave a gap for the coax connection. Smooth all cuts. Aluminum oxide sandpaper is best so that you do not leave residues of other metals on the aluminum. Clean the outer and inner edges of the tubes, since you will be inserting rods or tubes inside the main elements.

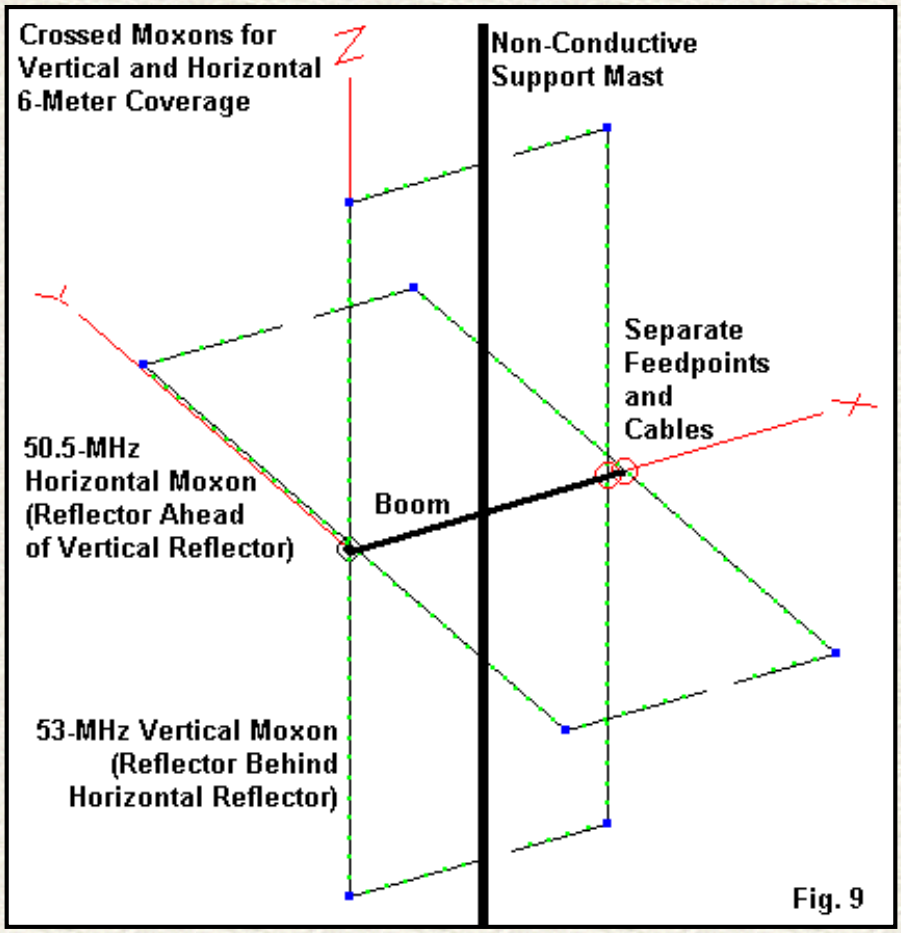
Now return to the assembly process and complete the Moxon rectangle. If you have used sufficient care, it should be right

on target on the first try. However, you can always trim a bit more from the reflector and driver tubes at their centers--and a little expansion that leaves a little inner reflector tube showing and widens the driver gap a small amount should do not harm.

I assume that you have a boom-to-mast plate and stainless steel U-bolt hardware. As well, I assume that the coax from the feedpoint has a connector and a double-female in-line connector. If those assumptions are correct, you are ready to put the Moxon rectangle into service.

Crossed Moxons

Suppose that you have both SSB and FM operations on 6 meters and that you decide that the Moxon has characteristics that are suitable for each job. Then you may wish to think about **Fig. 9**.



The outline sketches crossed Moxon rectangles. Since the low-end horizontal version is a bit larger than the vertical high-end version, let's place the horizontal reflector slightly ahead of the vertical reflector. Actually, it will make no difference if the two reflector join at the center. However, we shall need physically separate feedpoints--and separate feedlines as well. However, the assembly will fit on a single boom.

Crossing Moxons and operating them separately makes no difference at all to the performance of either one.

One precaution applies to either a single vertical Moxon or to crossed Moxons: we shall need a non-conductive boom if we attach the mast at the center of the boom. (6-meter arrays are heavy enough where I do not recommend extending the boom rearward for attachment to a mast without a further extension and counterweight.) Schedule 40 PVC has two sizes that nest reasonably well for stiffening the material for mast use: 1" inside 1-1/4" nominal (closer to 1.25" and 1.5" actual outside diameters). You will need about 3.5' to go from the boom to the edge of the vertical Moxon and perhaps another 2' above any metal structure (like a tower), plus a few more feet to reach a rotator. (An old TV rotator is more than sufficient to handle even the crossed Moxons). However, do not use more PVC mast than you need to do the job.

A Wire Moxon for 6 Meters?

As a final note, we should address the question of making a wire Moxon. A #12 or #14 copper wire Moxon is feasible for the horizontal version only. However, the thin wire will narrow the passband severely. If you operate within a very narrow spread of frequencies at the low end of the band, then you may consult one of the design aids and set up a wire version. It will perform well--as well a the tube version. As well, it may be easier to make sharp corners and trim to length.

However, a wire version of the vertical Moxon is likely to prove unsatisfactory for repeater hopping. It will work well for monitoring a single repeater--or a couple that are within a half-MHz of each other (allowing for the frequency split). However, for general coverage of the FM region of 6 meters, a version with fatter elements is strongly advised.

We can illustrate the opportunities and the limitatons of a wire Moxon for 6 meters with a simple example. Let's design a wire Moxon for horizontal use around the design frequency of 50.5 MHz. The dimensions will be as follows:

Dimensions for a 50.5-MHz Moxon Rectangle

El. Dia.	A	B	C	D	E
AWG #14 (0.0641")	84.86	12.53	2.61	15.95	31.09
0.25	84.31	11.65	3.44	16.08	31.18

I left the figures of the quarter-inch version for comparison. As shown in **Fig. 10**, the wire Moxon has a steeper gain curve and a sharper front-to-back curve than the tube version of the same antenna. However, for local and net operations, these figures may be very adequate.

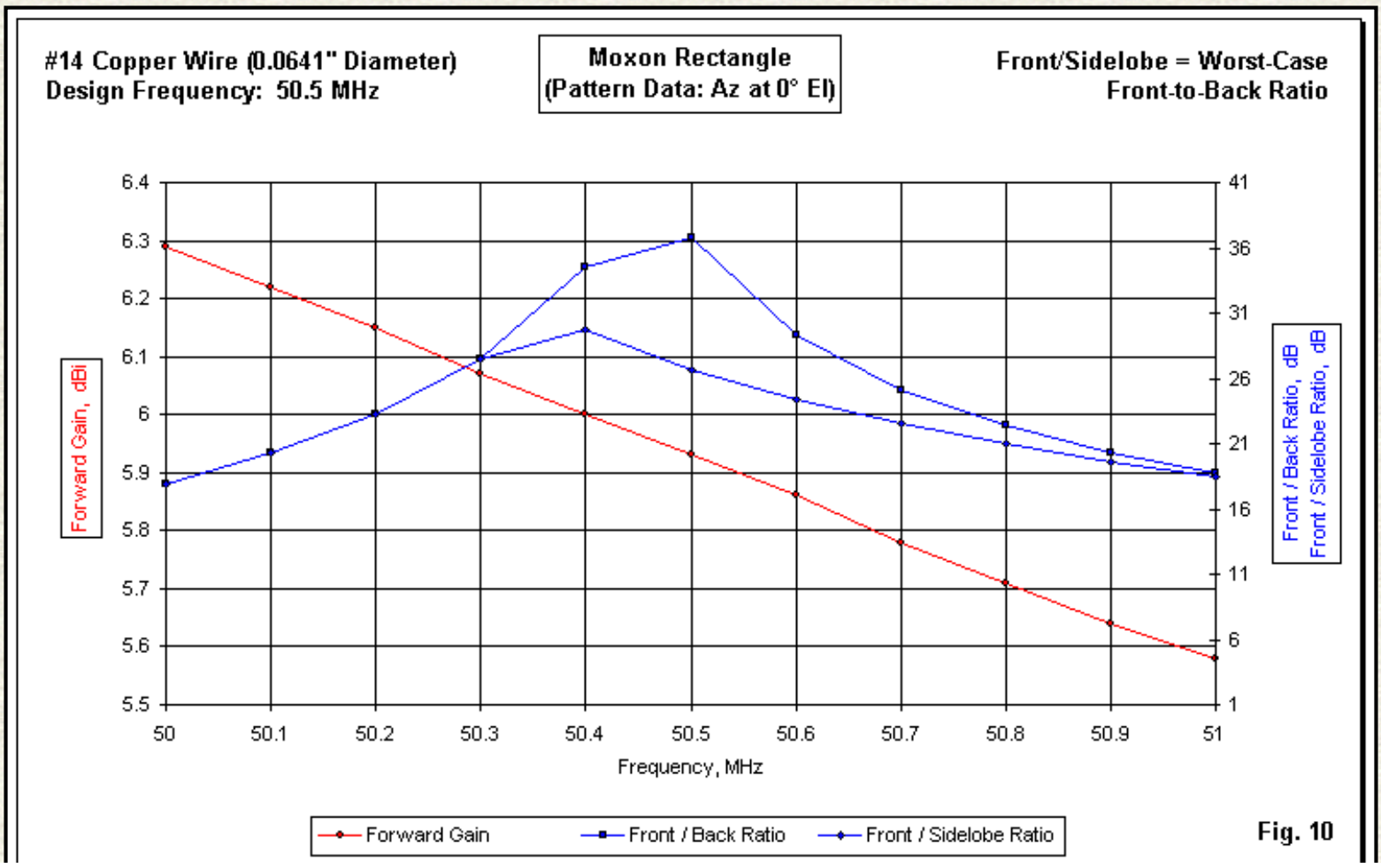


Fig. 10

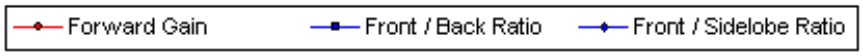


Fig. 10

The SWR curve, while steepr than the ones for the tube versions of the antenna, guarantees coverage of the first MHz of the band. See **Fig. 11**. You may note in the two graphs that the front-to-back ratio peaks just below the design frequency, as does the 50-Ohm SWR curve.

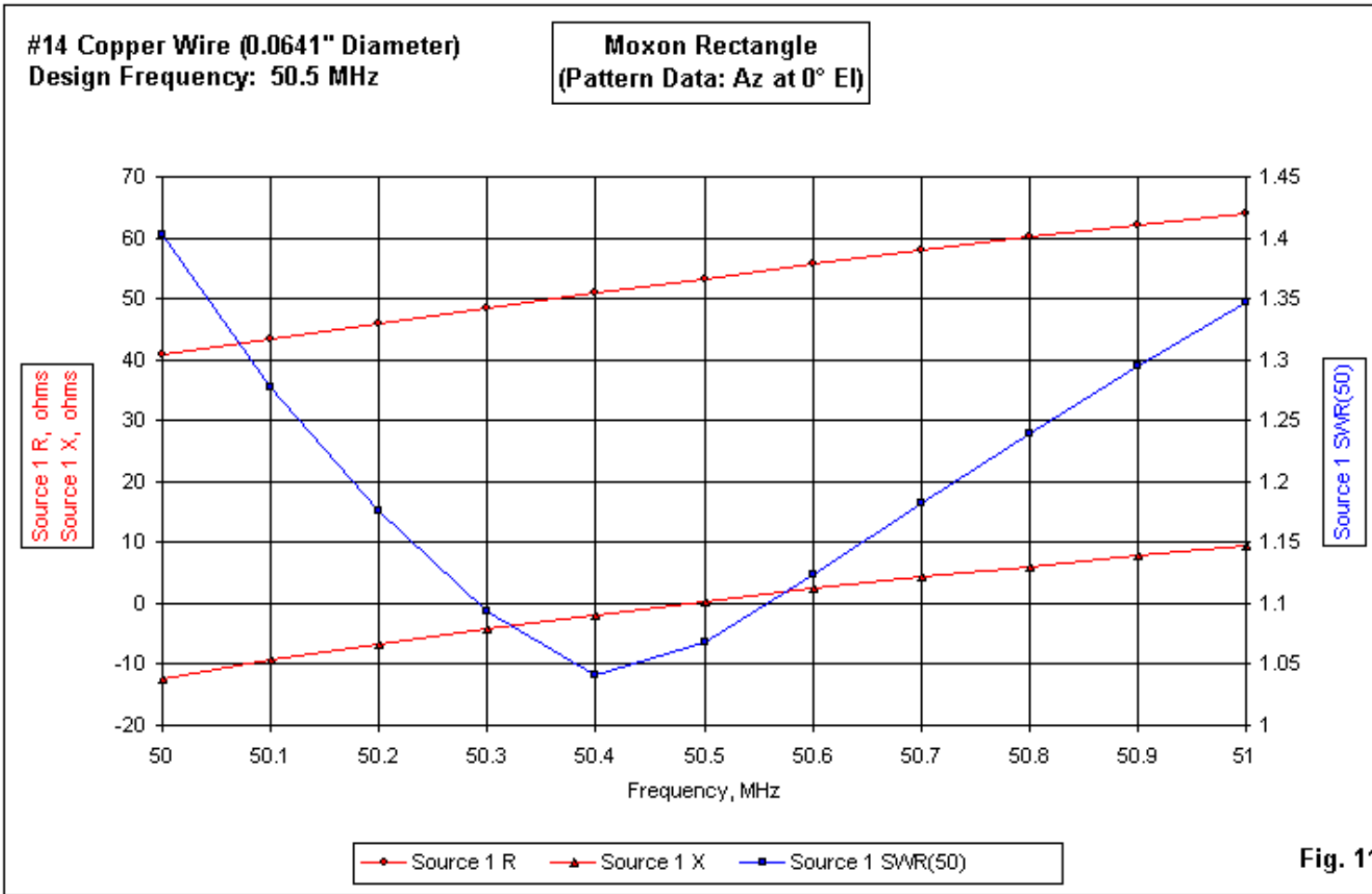
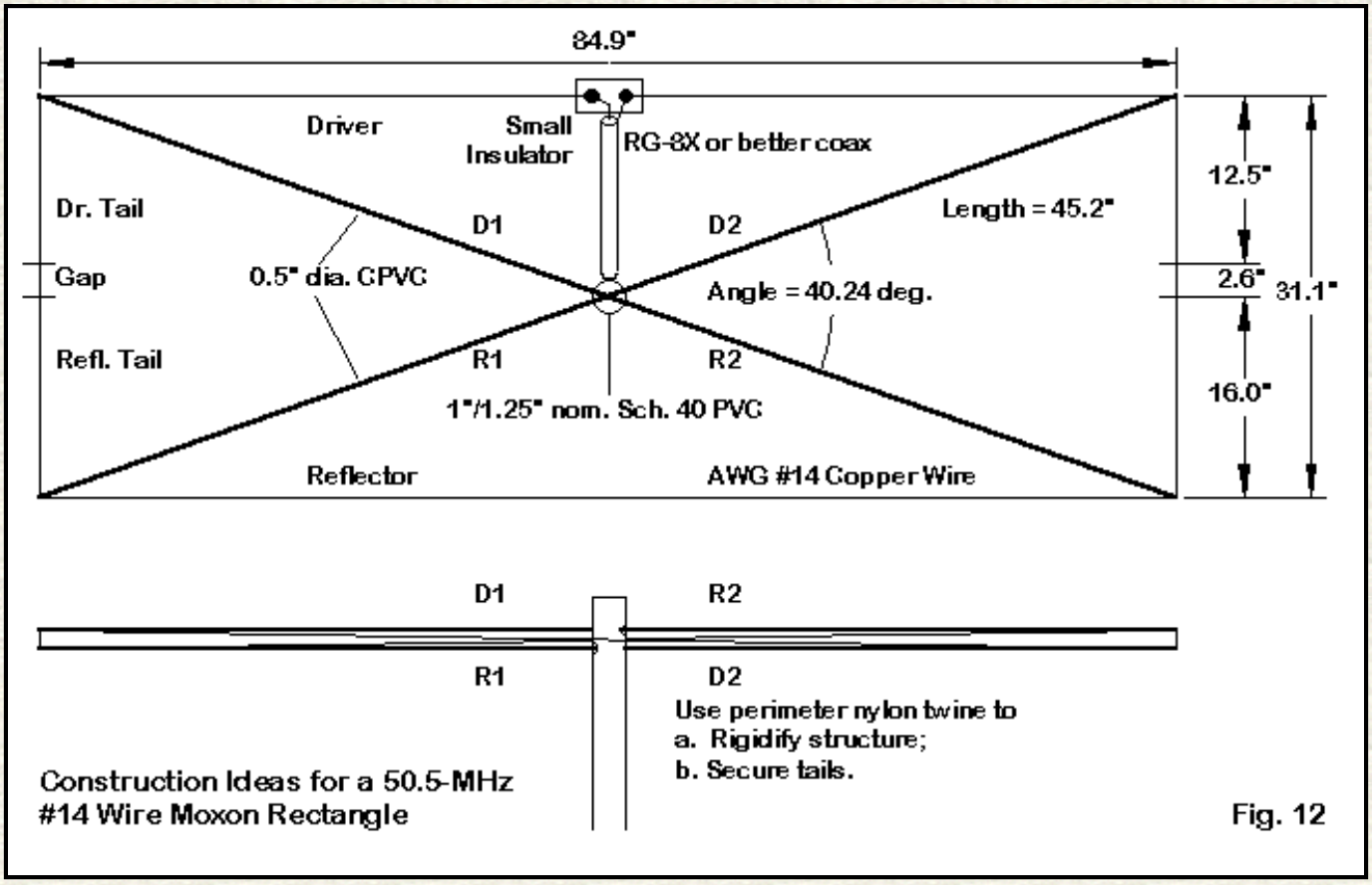


Fig. 11

You can make the frame from 1/2" or 3/4" CPVC. By passing the frame through carefully aligned holes in a central Schedule 40 PVC mast, you can cement the structure together. Each cross arm will need to be just over 7.5' long (3.75' each side of the mast). As well, you will have to plan your angles for the holes carefully to get the correct shape between corners. However, if you are only a little off, you can stress the arms with Nylon line (1/8" to 3/16") to perfect the shape of the ultimate support structure. If you adjust the holes in the mast, then add through bolts to finalize the positions of the support arms. The details of a suggested construction for the frame, line, and wires appears in **Fig. 12**.



Even if you you get the angles between supports correct, you may still run a length of nylon or similar line from the corners along the line of the tails. Then, tape the tails to this line, and the ends (raw cut and not looped) will stay in alignment and maintain their spacing. (I tend to prefer to use a full perimeter line to pre-stress the frame so that it maintains its shape under all conditions.) The resulting wire Moxon very likely will be considerably cheaper then any of the tube versions, since we can make it from PVC and household wire, along with a little hardware at the feedpoint.

Whether the Moxon is the right antenna--and which version is the one to build--depends on your own analysis of operating needs. Do not build one unless it will do the job that you need. But if you do build one, build it carefully, and it will work correctly without further field adjustment.

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The VHF Beam/Yagi Antenna

ARRL Technical Information Service page · TIS Menu page

Additional information on this subject and related topics can be found in back issues of *QST* and the following:

The ARRL Antenna Book
Physical Design of Yagi Antennas

The ARRL has an extensive catalog of books and materials related to Amateur Radio.

Articles

Note: Some of the following articles are in Adobe Portable Document Format (PDF) files. To view and print these files, you'll need a copy of Adobe's Acrobat Reader program. (Version 3.0 or later required). More information here.

- Homebrewing a 6-Meter Yagi (from *QEX* RF column) MEMBERS ONLY
  QEX January/February 1998, p. 52-57
  In typical Zack Lau, W1VT fashion, the element and mounting clamps are machined, but good ol' hose clamps and hardware store mast mountings will work on this 4-element antenna.
- Small 70-cm Yagi (671,292 bytes, PDF file) MEMBERS ONLY
  QEX July/August 2001. pp. 55-59
  A 6-elements designed for a wide bandwidth.
- 2x3=6 (84,913 bytes, PDF file) MEMBERS ONLY
  QST February 2000, pp. 34-36
  Two three-element 6-meter Yagi designs.
- 7dB for 7 Bucks (844,127 bytes, PDF file)
  QST April 1993, pp. 54-55
  An inexpensive 3 element 2-meter Yagi made from PVC and welding rod.

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5-Band Worked All States Pin -- Colorful design. 1-inch diameter pin with self-locking "tack" backing.

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Worked All States Pin -- You won the award...now show off your accomplishment!

DXpeditioning--Behind the Scenes -- A manual for DXpeditioners and DXers, covering virtually all aspects..from

planning to operating, traveling, QSLing, and more!

- A Five-Element, 2-Meter Yagi for $20 (457,324 bytes, PDF file) MEMBERS ONLY
  QST July 1999, pp. 34-37
  Get about 10dB from a modified Radio Shack FM broadcast receiving antenna.
- Circular Polarization and OSCAR Communications (1,401,889 bytes, PDF file) MEMBERS ONLY
  QST May 1980, pp. 11-15
  OSCAR users are switching to circular polarization to lessen signal fading. Build this low-cost antenna system and hear what you've been missing.
- A Tri-Yagi for 50 MHz (862,492 bytes, PDF file) MEMBERS ONLY
  QST June 1980, pp. 14-15
  For the performance of a 6-element wide-spaced beam on a boom half its normal length, try this "trigonal" reflector scheme.
- An LPDA for 2 Meters Plus (839,246 bytes, PDF file) MEMBERS ONLY
  QST October 2001, pp. 42-46
  This Log-Periodic Dipole Array (beam) antenna covers 130-170 MHz for those who also like monitoring.

Note:

Contact information for suppliers mentioned in the above articles should first be confirmed using **TIS Address Database Search**.

Bibliography (Members Only)

ARRL Periodicals Index Search - This database contains the *QST* index from 1915 to the present and the *QEX* index from 1981 to the present. For *QST* issues from 1970 to the present, and some selected articles back to 1922 (when construction articles featuring tubes began in earnest), identifying keywords have been added to the technical articles. By entering keywords (ANTENNA) or combinations of keywords (CONSTRUCTION ANTENNA VERTICAL HF) into the **Title words:** field, you may create dynamic bibliographies.
Technical article KEYWORD list. Hints for more successful searching

Suggested keywords for more articles like the ones on this page are: CONSTRUCTION VHF BEAM ANTENNA

Web Links:

- The Tiny 2
  The "Tiny 2" is a great little 2 meter beam. It has some really interesting properties and it is a fantastic first time antenna project for the beginner.
- The Tiny 2+1
  Renamed the Tiny 3. As with any standard three element design, this three element direct connect antenna utilizes a reflector and a director to achieve gain.
- VHF / UHF Direct Connect Beams
  Here are some lengths and spacings for various direct connect beams.
- 2 Meter Antenna Design Gallery!
- Antenna Round Up
  Antennas for 6, 10 and 2 meters
- Stacking Yagi Antennas

- A Simple Seven Element Yagi Antenna
  A JavaScript that quickly calculates the dimensions for a seven element Yagi Antenna. Simply enter the frequency in Megahertz and the script will do the rest.

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A Six element antenna for the 50 MHz Amateur Radio Band / Version 4

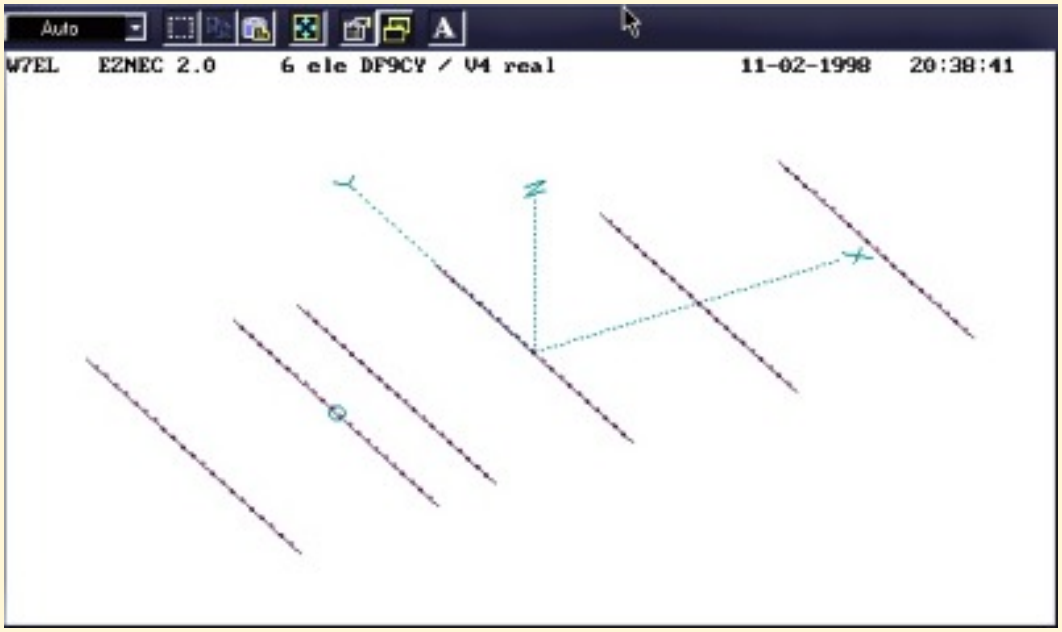
© Christoph Petermann DF9CY 1998

Last Update: 31 May 2001

A better view of the 6 element

I have mounted the 6 Element antenna about 2 meters above my 3 element antenna for 28 MHz. (The 28 MHz antenna is a well performing "EZNEC redesigned" CB radio antenna)

This plot shows the antenna "as seen" by the computer.



Here is a short description of my new SIX element antenna.
Design Frequency: 50.150 MHz

- **Usable Bandwidth: 500 kHz**
- **Return loss: better than 10 dB within 500 KHz with 22 dB @50.150 MHz.**
- **Gain: > 11.2 dBi**
- **Front/Back ratio: > 12 dB**
- **Vertical sidelobe suppression: > 12 dB**
- **Horizontal sidelobe suppression: > 19 dB**

Version List

- **Version 1:** Only a simulation version.
- **Version 2:** Has been published here. Version 2 has slightly changed element lengths and positions over Version 1. Gain has been reduced to 11.3 dBi, but it has an even better bandwidth performance. The first antenna of this type (Version 2) has been built by **Hellmuth, DF7VX.** It seems to work reasonably well. The design goal's resonance frequency has been hit exactly. Hellmuth is using this antenna since the 1998 Es season and worked a good number of DX stations with it. I have built a 2nd antenna and tried out a number of different feeds with it. This included folded 200 Ohms dipoles and Gamma matches.
- **Version 3:** This version was published here in July 1998. It has been built several times in England and America with good success.
- **Version 4:** The previous version suffered from having the resonance frequency a little low. I made some minor changes. This antenna is ready in operation here now and I am delighted with its performance in 12m above ground. It has a clean pattern with a good front/back ratio. SWR is excellent and broadband at least over the DX portion of the 6m band. It outperforms well the 4

element I have described here also.

- **N7BW Realisation**Here is a version built by John N7BW. The **6-ELEMENT** and the **FEED-POINT**, at the driven element with an open dipole. (Images)

ELEMENT #	POSITION	center-to-end element length	Diameter
	Meters	*Meters*	*Millimeters*
Reflector	0	1.450	12
Dipole	1.114	1.385	12 folded dipole
1 st Director	1.585	1.328	12
2 nd Director	2.657	1.328	12
3 rd Director	3.942	1.327	12
4 th Director	5.327	1.320	12

Driven Element

- I strongly recommend building a folded dipole. I did have VERY poor experiences with Gamma Matches in the past. They are for low impedance antennas only and they provide the necessary transformation. So they are narrow banded and if you have not the high quality material at your hand, you will sacrificy some of your gain you have made the antenna for.
- A folded dipole needs a balun: take a good cable like RG400 teflon for it. The velocity factor of this cable type is 0.82; a length of 2.45m will do the job. It is expensive, but you want the DX, don't you?
- If you do not want to build a folded dipole, make an open dipole mounted insulated from the boom. A 1:1 RG400 transformer is built within 10 minutes and -voila- you have your antenna going.
- I designed the antenna for 50 Ohm (i.e. 200 Ohms if you have a folded dipole).
- TIP: Buy the complete folded dipole from **KONNI Antennen in Esselbach**

Elements

The diameter of the elements is 12 mm AND NOTHING ELSE. The antenna will not work, if you take something else. (J'espère, que c'est O.K., Jaques ?) The elements are mounted insulated above the boom. I have about 4 mm space from the boom to the elements. I use industrial spacers mounted with a 4 mm metal (!) screw centric to the 20*20mm boom. The deviation from an element being really isolated is minimal. Indeed I bought the whole element at **KONNI Antennen**. Buy reflector elements only and cut them to the given length.

BOOM

This antenna is calculated for a 20*20mm of 2 mm strength boom which I made from material from a metal shop and some scrap from Tonna 2m Yagis ...
I replaced the support boom you see on the image by a longer one. This gives more strength. In February

1999 the antenna survived a monstrous wind storm with gales up to 165 km/h.

Interactions
If you put this antenna on top (or below) your existing antennas: be aware that you reduce the antennas' performances. Keep this antenna at least 3.0 m away from ANY shortwave antenna and about 1.5m from other VHF antennas.

Stacking
If you are intended to stack 2 antennas, space them minimum of 1/2 wavelength. At half wavelength, the existing vertical sidelobes wil DISAPPEAR with a gain around +2.1 dB. Stacking the antennas at 1 wavelength increases stacking gain to about 2.7 dB and sidelobes go up. I would like to space 2 antennas at 1/2 wavelength, but the space on the tower is limited.

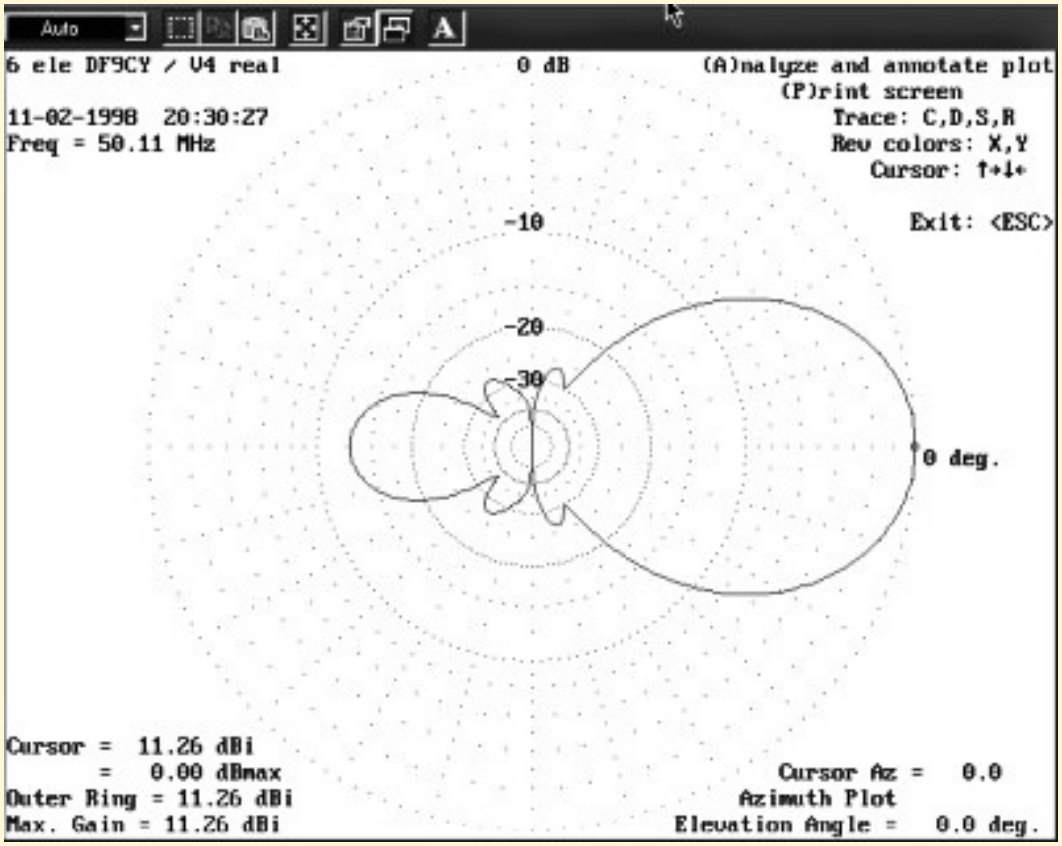
Performance
I have been reported and I can confirm myself: That antenna works. I see a clear pattern with the antenna up at 12m. There is almost no sidelobe and the F/B ratio is good. The SWR is also as expected.
It clearly outperforms the 4 element I did have up here through 1998. It is not more than 2.5 dB (or half a S-unit), but having more energy at the horizon gives you an extra advantage. The OZ7IGY beacon at 200 km line of sight is up about 2 S-units now (I am wondering myself ...) and the other beacon OZ6UHF at 300 km is also better now.
After a few months I can say, the antenna does a very good job. Tropo signals are often audiable over distances of more than 400 km.

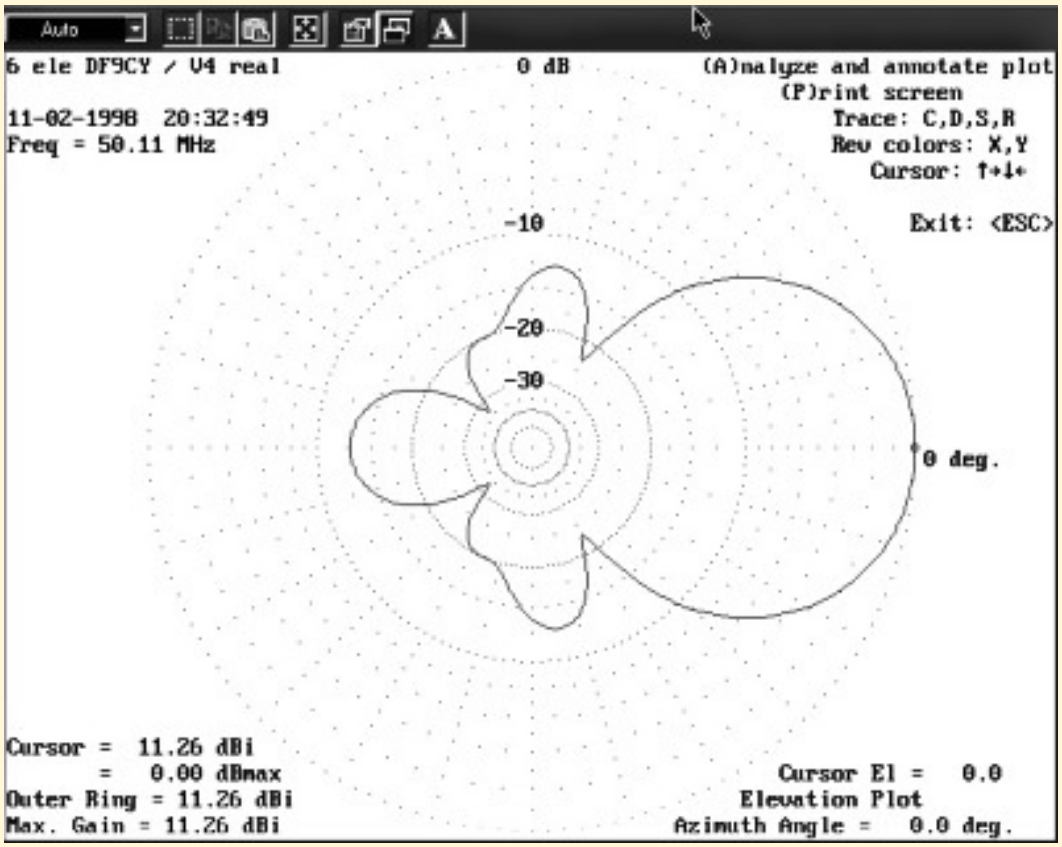
After 2-1/2 years ... (May 2001)
I am still very pleased with the performance of the antenna. It withstood all brute gale winds here in my "seaside-resort". But 30 black birds (ravens) destroyed the antenna completely already twice. Now I have an 8 mm aluminium tube in the centre of on element strengthening against the weight of these birds. DX performance is better than everything I had before. I like the clean pattern. Certainly the front/back ratio of 12 .. 15 dB could be better, but you can have this by giving up the clean pattern and some gain. I know of about at least 12 hams having built this antenna and so far all are quite pleased with it. I would like to have a stacked version of two at a spacing of 4m, but my antenna tower is somewhat weak for that. My feedline is a 22m run of AirCell cable. I get a strong enhancement in noise when I turn the antenna across the galactic centre in summer. I had a low noise GaAs-Fet amp at the antenna for a while, but this was not necessary; so it is down again. In October 2000 I moved the antenna from 10.5m to 13m above ground level. The patterns became even cleaner and I succeeded in nice F2 QSOs to Far-East and the U.S.. The nicest thing however is to work the DX with a flawlessly well-performing homemade antenna.

EZNEC
I provide you with the EZNEC simulation. So download the EZNEC-FILE: **50-6cy4r.zip**

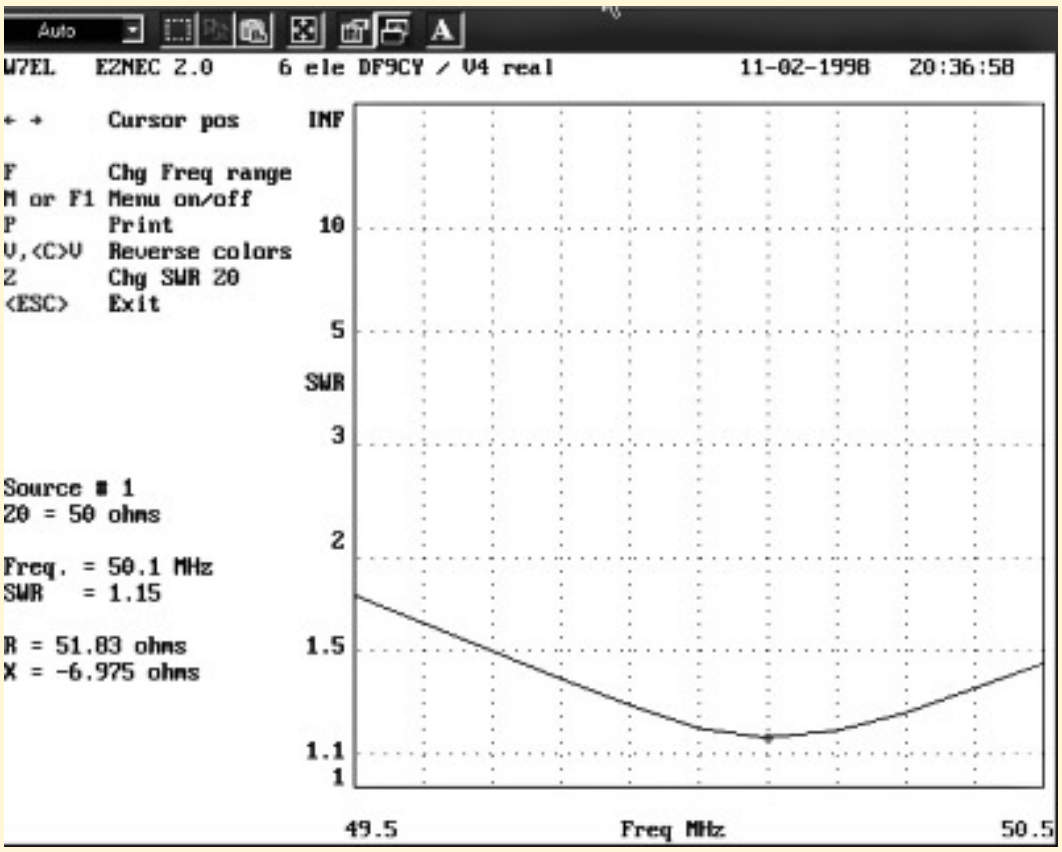
DIAGRAMS AND SIMULATIONS

Below are the diagrams for the azimuth plane (upper) and the elevation plane (lower). Sidelobes are barely visible in the azimuth plane and well 12 dB below maximum gain in the elevation plane. These are the free space simulations.

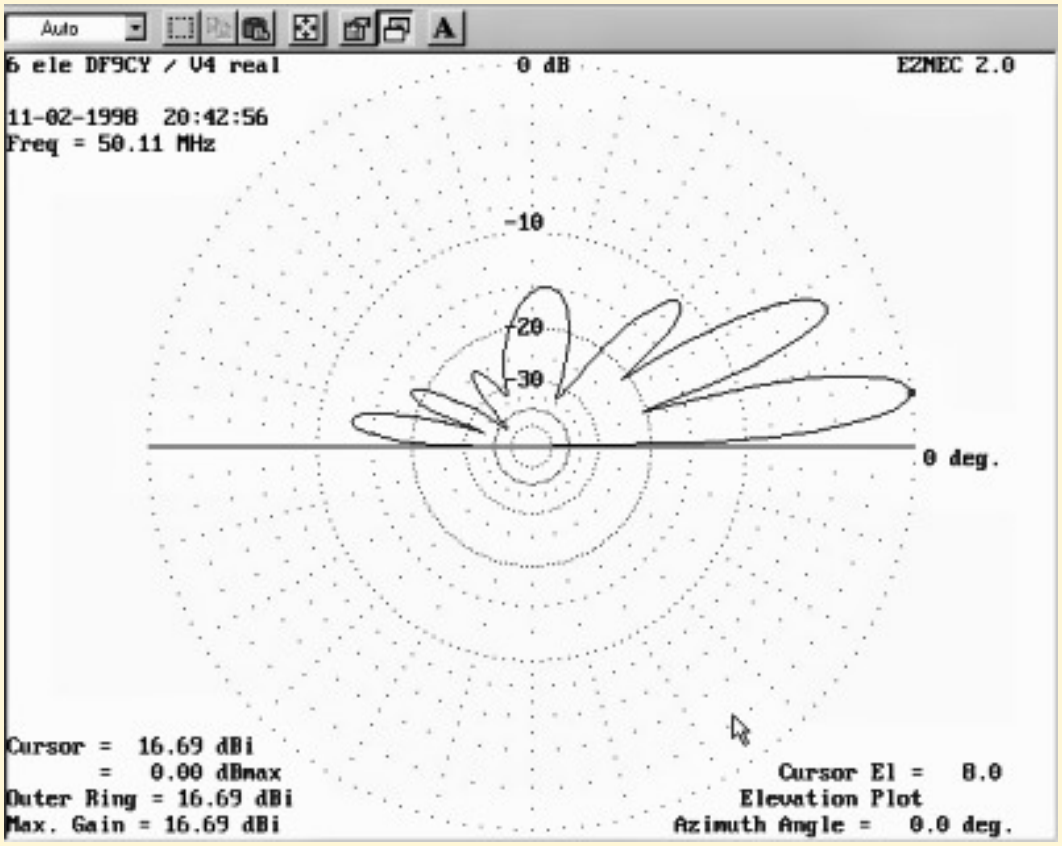




The SWR plot shows the optimum SWR at 50.110 MHz



This simulation shows the elevation pattern over real ground with the antenna mounted at 10 meters above ground.



My special thanks go to Roy Lewallen W7EL for programming this fantastic *EZNEC 2.0*

If you want to have more background knowledge on antennas and how-to make them, you will find an ultimate resource here:

W4RNL
L.B.Cebik W4RNL ultimate ANTENNA Site. A must !

Mail any comments to: df9cy.petermann@t-online.de

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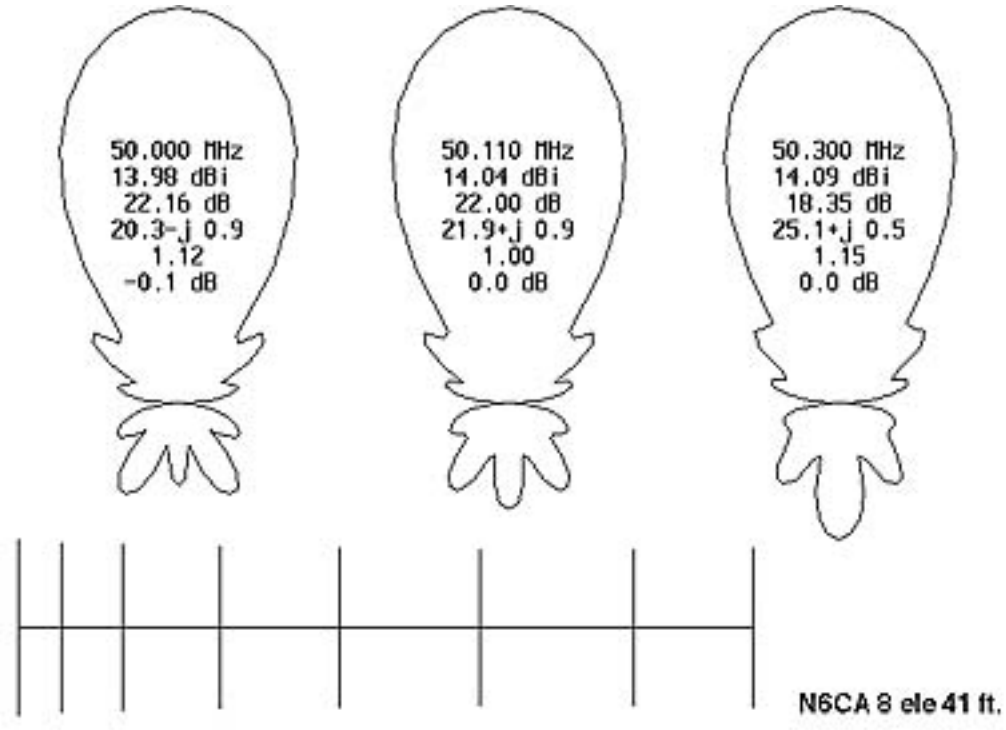
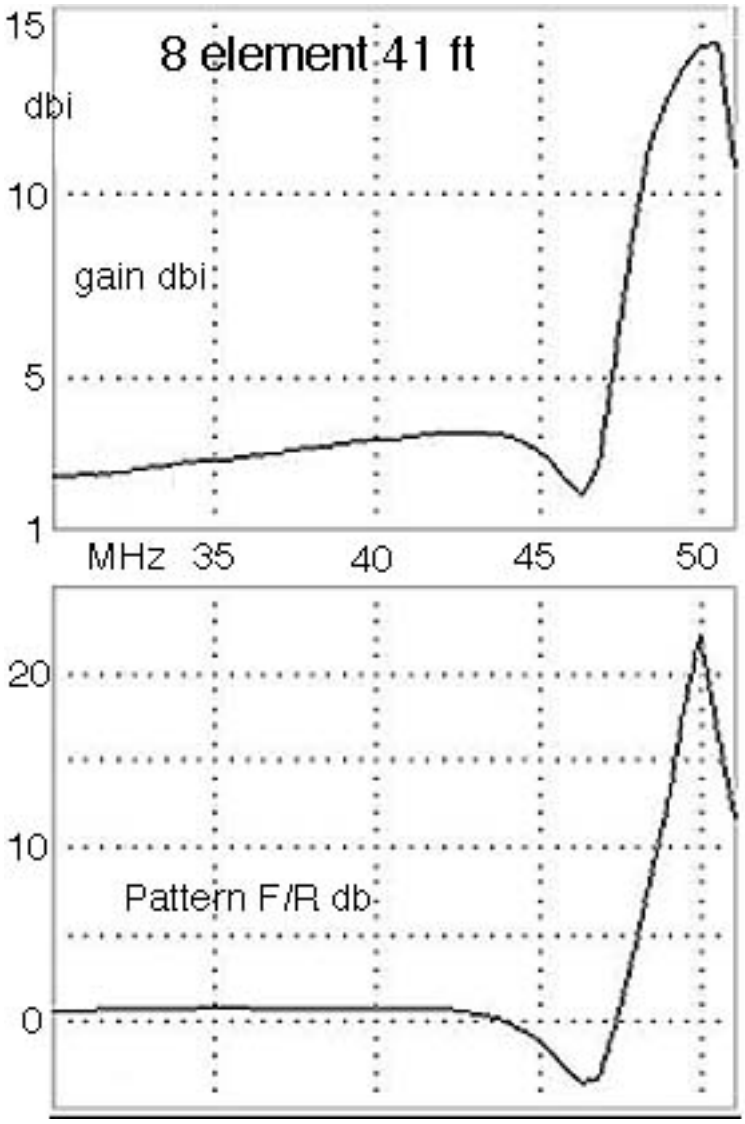
GO (back) and visit my **homepage**

Antennas: (updated Feb 25, 2002)

8 element, 41 foot boom, mounted at 87 feet. 14.0 dBi gain, homebrew design (pictured on homepage) used for F2 & multi-hop Es: 1.75" diameter boom,center section has .125 wall, all others .062 wall..250" dia elements thru the boom insulated, add 8.7% boom diameter to element length for this boom. Tee match, 1.150" spacing, .250" diameter bars, length = 11.5 inches each side, fed with half wave length coaxial cable balun from UT250 with shorted and grounded quarter wave length stubs each side of Tee match. The grounded quarter wavelength stubs did absolutely nothing for suppressing rain static nor summer time "dry air static" common to this area. Will leave them out next time. Driven element dimensions may be too short as I forgot to re-measure it after final adjustments were made....will do that next time down.So far this cycle this antenna has performed extremely well on the F2 and TEP paths. Its clean pattern has paid off on suppressing noise and local QRM while looking for DX.

Element	reflector	driven ele.	dir #1	dir #2	dir #3	dir #4	dir #5	dir #6
Length	116.276	112.600	109.534	107.556	105.776	103.754	102.700	104.754
Spacing	0	28.530	69.265	134.597	213.310	306.557	408.697	488.066





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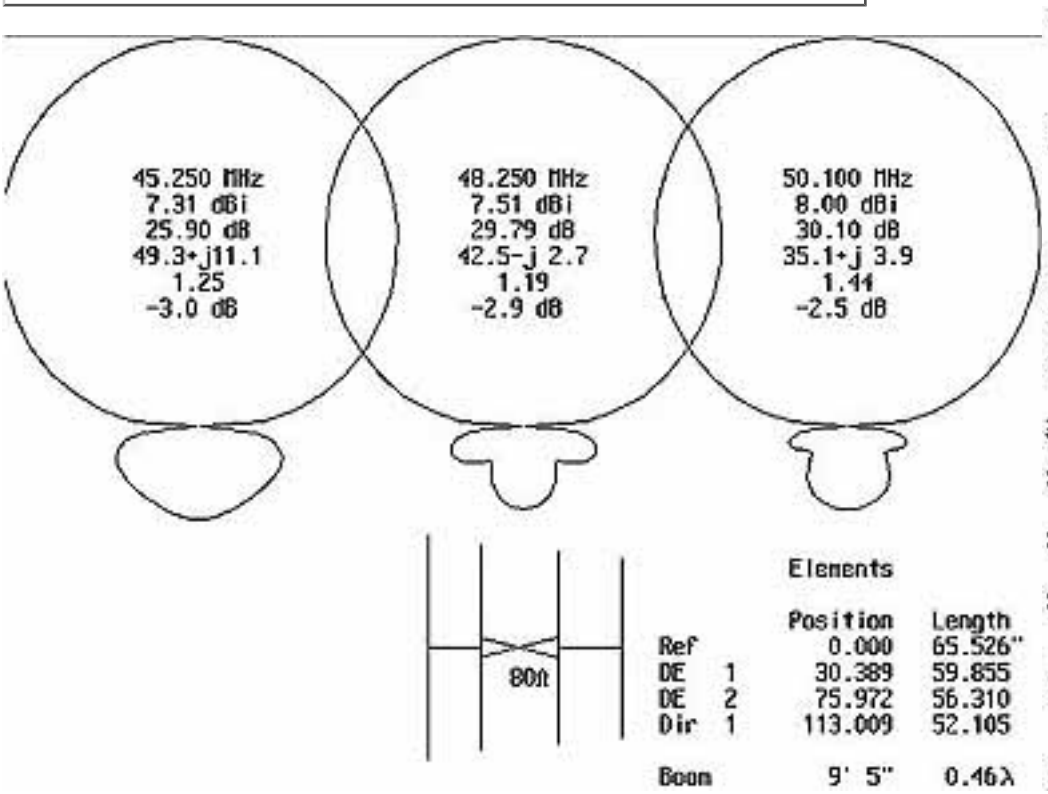
4 element broad band yagi

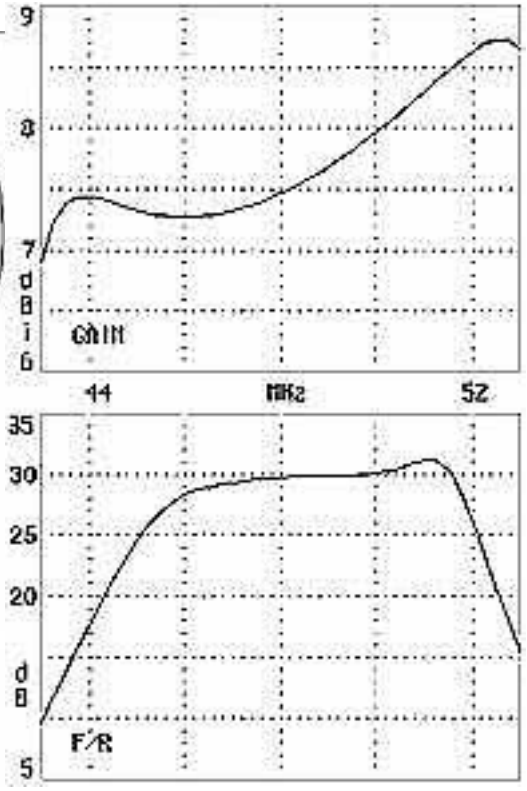
This antenna was the final choice for the all purpose noise cancelling/MUF antenna. It has a super clean pattern yet and enough gain to be useful. Too much gain would make it harder to use in noise cancelling applications. It is designed for many uses: 1) noise souce antenna for the noise cancelling receiver. 2) a 43 to 51 MHz antenna for observing MUF as the band opens up. 3) as a secondary wider beam width six meter antenna. All of the aforementioned points require an antenna design with a good pattern, resonable gain and good impedance.

Compare the gain-bandwidth of this antenna to the rather narrow gain-bandwidth of the 8 element yagi above. Many MUF indicators are between 40 and 50 MHz. Mounting tests will be be performed first to assure there will be adequate isolation between the 8 element yagi and this one. I do not want any deterioration of the 8 element yagi's performance. They will be 3/4 wavelength apart (15 feet). There should be around 20 db isolation between the two antennas (17 db measured...yuk!). I'm still amazed at the computer performance of this antenna considering the bandwidth....

The boom is 9 foot 7 inch, long 1 inch OD 0.065 wall. Elements are .250 inch diameter alu tubing as usual and are thru the boom and insulated. The driven elements are split in the middle and mounted on Delrin blocks. The phasing/matching lines made from UT141 semirigid coax (only outer conductor used). Impedance of the open wire line isn't that critical and can run anywhere from 70 to 100 Ohms without much effect. Add 5.1% of the boom diameter for thru the boom insulated compensation to the element lengths for this boom. Results of testing will be posted when I get them. Antenna seems to perform as expected. Better signal to noise ratio that big yagi at lower freqs and of course great directivity. Impedance is as computer predicted at <2:1 due to higher impedance openwire feeder to eliminate water accumulating between the conductors. The output of this antenna is split or switched to drive the filter/preamplifier/R8500 and the noise cancelling receiver. It's a great ant with multiple uses for us one tower guys. A separate low power T-R relay is used to isolate the output of this antenna when transmitting on the 8 element yagi.

Element	reflector	driven ele, #1	driven ele. #2	dir #1
Length	131.05	119.7	112.6	104.2
Spacing	0	30.39	75.97	113.0





Open wire feeder made from UT141 semidrigid coax, only the outer conductor is used, spaced .150 inches. This is greater impedance than the 80 Ohms so it won't collect water between the conductors. It is important to maintain the air dielectric. Delrin spacers are used throughout. Balun is ferrite beads on outside of coax held in place with shrink tubing.

---

6 element , 24 foot boom, 12.1 dBi design used for Es: Units=inches, .250" dia elements thru the boom insulated, 1.0" diameter with.125 wall center section and all others .062 wall. Add 5.1% boom diameter to element length.Tee match, 1.150" spacing, .250" diameter bars, length = 14 inches each side, fed with half wave length coaxial cable balun, RG142.

Element	reflector	driven ele.	dir #1	dir #2	dir #3	dir #4
Length	116.227	113.50	109.720	107.170	105.708	103.436
Spacing	0	40.5	80	142	216	287

4 element , 15 foot boom, 10.8 dbi gain, designed for light weight portable antenna. Clean pattern: all lobes >-21 db. Usually run on 21 foot mast. Boom = 1.0 inch diameter with .062 wall. .250 diameter elements with .049 wall. Thru the boom insulated elements in table above already have add 5.1% diameter of boom (.050 inches) added to element lengths.Tee match: .250 diameter rods, 1.0 inch spacing, 12.75 inch length on each side from boom center with 1.625 inches protruding beyond shorts. Shorts are made from 1/4 x 1/2 inch copper bar. RG142 half wave length coaxial balun. Antenna picks up 1 db gain over the shorter boom 12 footer. This one built with splice at the driven element for easy assembly while portable.

Element	reflector	driven ele.	dir#1	dir#2
Length	115.4	112.55	107.6	106.2
Spacing	0	47.7	108.5	177.7

3 element, 5.2 foot boom, designed for super clean pattern (side lobes> -29 dBc) and low gain! Antenna used for 50 MHz noise cancelling receiver. Half wavelength coaxial balun, RG -142. Units=inches, .250" dia elements thru the boom insulated, 1.0" diameter boom, add 5.1% boom diameter to element length. Tee match 1.0"spacing, .250" diameter bars, length = 11 inches each side, fed with half wave length coaxial cable balun.

Element	reflector	driven ele.	dir #1
Length	118..0	114.2	106.5
Spacing	0	24	62

All antennas were designed with YO, Yagi Optimizer.

Radio:

Homebrew hi level transverter, Rx filtered for 50 to 51 MHz pass band, -45 dBc at 54 MHz, noise cancelling input, adjustable phase & gain. 22 dB conversion gain, 2.5 dB noise figure. Terminating low pass filters on Tx.

Incorporating a FT1000D modified for use as the primary 50 MHz IF radio. TS930 had more than 60 thousand hours of six meter operation on it. (tired). First IF filters need to be added to FT1000D.and similar to 930 transverter mods need to be incorporated but I am finally using it. Going to add a new 8 pole first IF filter, PIN atten and hi level bipolar first IF amplifier replacing the original dual gate mosfet aa 73 MHz.

Icom R8500 receiver used for band segment scanning and third receive frequency. Adding a 43 to 51 MHz "MUF" antenna (above) to drive a 5 pole band pass filter and hi level preamplifier and splitter which will become the "extra" input for several uses in near future. Design will appear under application notes.

3CX800 amplifier, homebrew, 1200 Watts (+60.8 dbm +/- 0.02 db) output at 1 dB output compression point (CW mode). Modifying a TL922A amplifier for "instant on" six meter amplifier.

90 feet of 7/8 inch hardline & 85 feet of 0.5 inch hardline to antenna feed point

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Optimised Six-Metre Yagi

by Brian Beezley, K6STI, ***Issue 39, October 1993***

Here's a design for a 5-element beam on a 23-foot boom with an unusual combination of performance characteristics. This Yagi comes within 0.2dB of the maximum gain possible on the boom length while keeping all back lobes 20dB down. This performance combination is very rare. The trick was to optimise the design over a narrow frequency range, 50.000 to 50.250MHz. Many 6-meter beams, both homebrew and commercial, are designed to work to 51 or even 52MHz. These designs invariably sacrifice forward gain and pattern quality for wide SWR bandwidth. When optimising a design over a narrow bandwidth, fewer elements are needed. As long as you have a certain minimum number, Yagi forward gain is determined by boom length, not element count. Elements added to the interior of this design won't increase its gain. (They may increase the bandwidth over which the pattern and SWR remain good, but this design adequately covers the low end of 6 meters.)

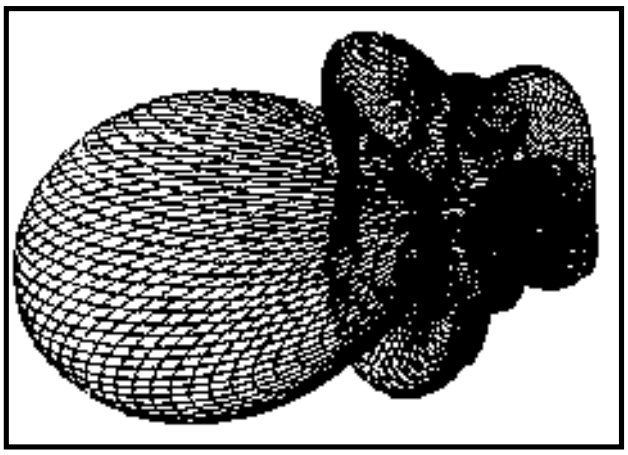


Figure 1 - 3D plot - Peak 10.26 dBd @ 50.110MHz

The free-space forward gain of this Yagi varies from 10.2 to 10.4dBd over 50.000 to 50.250MHz. These figures include conductivity losses of 0.08dB for 6061-T6 aluminium elements. When matched at 50.135MHz, SWR is less than 1.4 across the frequency range. The worst-case back lobe is 20dB down at 50.000, 21dB at 50.100, and rises to 16 dB at 50.250MHz. This design was simultaneously optimised for maximum forward gain, minimum worst case back lobes, and adequate and impedance. Conventional F/B was not optimised. This parameter takes the rear pattern into account at a single point.

The tuning of Yagi elements depends not only on their length but also on diameter, diameter tapering, and mounting method. These factors affect element self-impedance and thus alter antenna response. Thinner elements, tapered elements, conductive mounting brackets, and through-the-boom mounting shorten effective element length. Design dimensions are given for insulated, untapered, 0.375" diameter elements, for insulated, untapered, 0.5" diameter elements, and for elements with Cushcraft A50-6S tapering and mounting. If you use different elements or mounting methods, you'll need to adjust element lengths for optimum performance.

This design has a special property which makes it easy to adjust element tuning experimentally. The azimuth pattern has three back lobes which are equal in amplitude

only at 50.100 MHz. Since rear-lobe amplitude changes rapidly with frequency, you can easily verify that your antenna is correctly tuned. If the rear lobe is larger than the other two back lobes, the effective length of your elements is too long. If smaller, your elements are too short. You can use this simple test to obtain correct electrical behaviour for any physical element mounted by any method. When adjusting parasitic-element lengths, make equal changes to all parasitics. The input impedance of this design is about 12.5 Ohms. You can use any matching method as long as you observe the fundamental rule of Yagi matching: Never alter parasitic-element lengths or spacings of an optimized design to get a good match. Driven-element length has virtually no effect on gain or pattern, so you're free to adjust this dimension when matching. Don't after element spacing. Change parasitic-element lengths only to move the equal-back lobe frequency to 50.100 MHz.

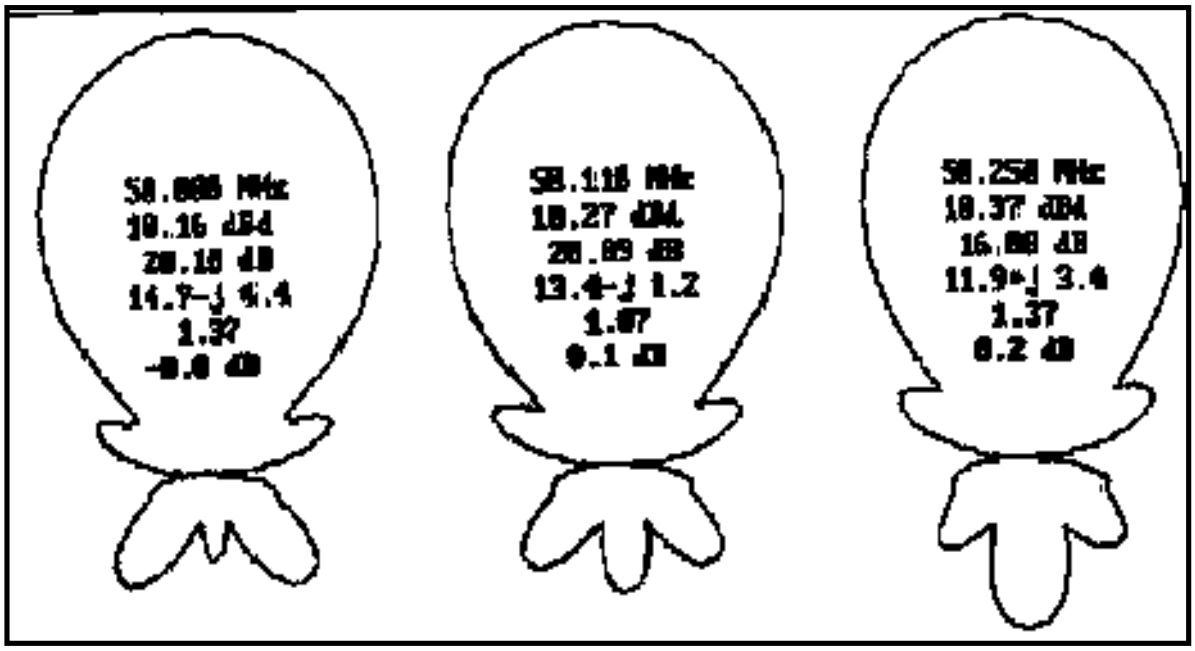


Figure 2 - Polar Plots @ 50.088, 50.110, and 50.250MHz

When I built this antenna in 1989, I gamma-matched it. However, I wouldn't do this with the knowledge I have today. A gamma match can induce current on the shield of a coaxial feed line. It can also induce current in the boom unless the driven element is insulated. These stray currents can reduce forward gain and degrade the pattern. You may be lucky as I was and get away with gamma matching, but why take a chance? Use a hairpin, T-match, or folded dipole and a good balun. The 12.5 Ohm input impedance transforms to a feed impedance of 50 Ohms for an equal diameter folded dipole. You can feed the folded dipole directly with 50 Ohm coax if you use a current type balun. You can make one by slipping ferrite beads over the coax or you can simply coil the coax into a few turns near the feed point. This design was developed in 1989 but was not published for some time. A carefully constructed, untapered, insulated-element version of this design came out 220KHz low. (I had to DF cordless phones below the 6-meter band to find the equal-back lobe spot!) I thought that perhaps some obscure environmental factor was responsible for the discrepancy (like the conductivity or dielectric constant of my composition roof). In typical ham fashion, I simply cut a quarter inch off each element tip and used the beam

successfully. But the 220-KHz anomaly continued to bother me. It wasn't until I began to use the sophisticated Numerical Electromagnetics Code (NEC) that I finally understood what was going on. NEC predicted that the antenna would perform as originally measured. The MININEC-calibrated algorithm I used to optimise this design had a built-in frequency offset! I immediately recalibrated all of my antenna-design programs to NEC. I subsequently found references to the MININEC frequency offset in the professional literature. If you optimise this Yagi for maximum forward gain without regard to pattern, impedance, bandwidth, construction tolerances, or reason, you can squeeze 0.2dB more out of the design. However, the back lobe degrades to just 11dB down and input impedance nose-dives below 5 Ohms. You say that you can match anything? That you never leave 50.110? That you don't experience rear-signal QRM on 6 meters, and that you want all the gain you can possibly get? What about those damn power leaks that always seem to start up from every direction whenever the band opens? What about the 3CO DXpedition which shows up on 50.300 and gets chased off the island before you're done retweaking your match up on the tower? If you're greedy about Yagi forward gain, you'll live to regret it.

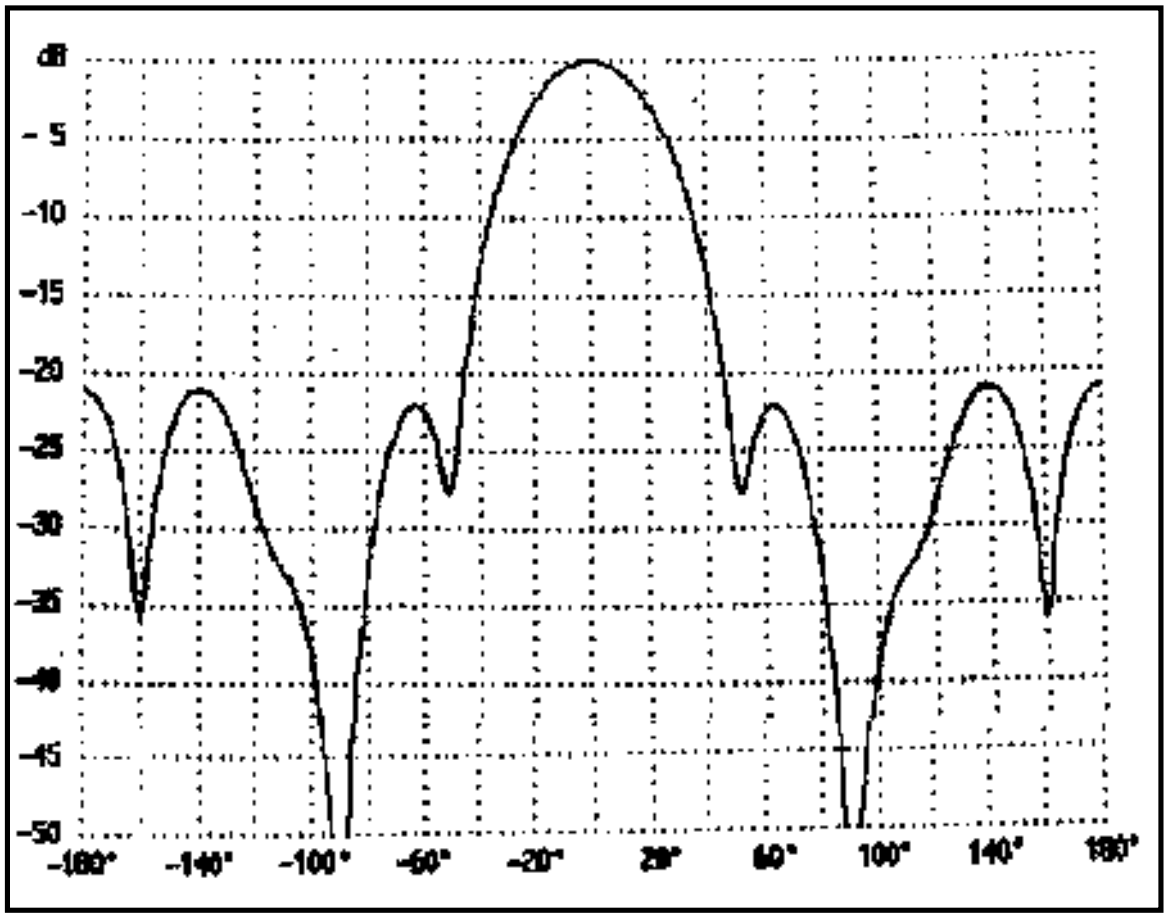
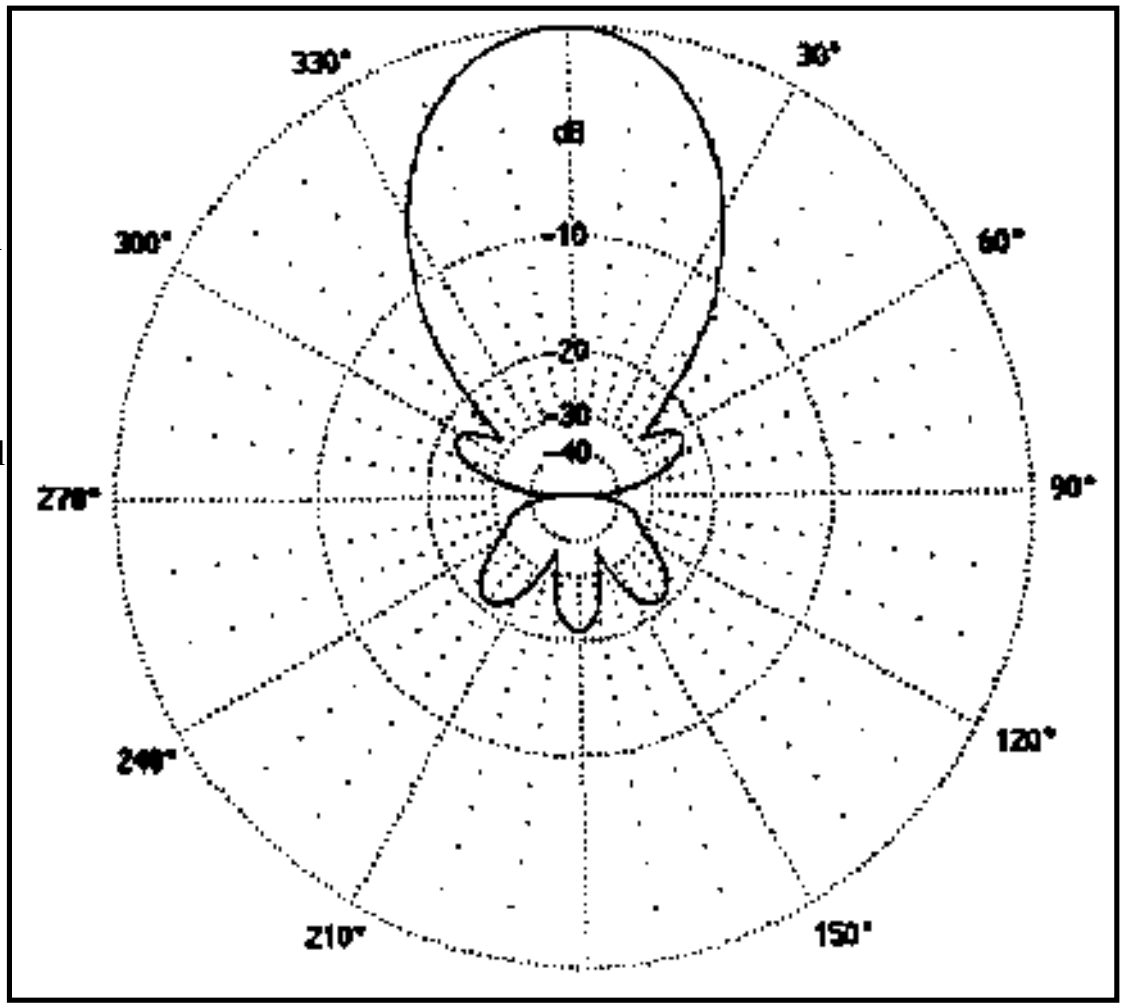


Figure 3 - Polar Plot - 50.110. 0dB = 10.26dBd

You can stack two of these Yagis for more gain. An H-plane stacking distance of 27 feet provides 3.1dB additional gain in free space. (Other stacking distances shift the equal-back lobe frequency away from 50.100MHz and require element readjustments However, unless the array is very high, you won't come close to free-space stacking gain in practice. The elevation patterns of Yagis at different heights don't combine favourably. For example, adding a second Yagi 27 feet below one at 50 feet improves gain less than 1.2 dB at elevation angles below 5 degrees. E-plane stacking is an attractive alternative. If

you space the booms 29 feet horizontally, you'll get 3.0dB gain over a single Yagi regardless of height. The 3-dB beam width will be 9 degrees, with deep nulls at 20 degrees and side lobes 9.4 dB down at 31 degrees. Before you go to the trouble of E-plane stacking, think carefully about the operational inconvenience of such a narrow main lobe. The cross boom must be non-conducting near the Yagis. If you'll settle for an improvement of 2.6dB, the 8-element Yagi listed below is much more manageable than a side-by-side pair of 5-element designs. To give you idea of how this Yagi compares with other designs, tables 1 and 2 show some NEC results at 50.110 MHz:

Figure 4 - Polar Plot - 50.110MHz



A50-5S, A50-6S, and 617-6B are wide-band Cushcraft designs. 6M2WL and 6M2.5WL are M2 Enterprises designs. NBS-5 and NBS-6 are National Bureau of Standards designs empirically optimised for maximum forward gain. Five is the subject of this article, while Four, Six, Seven, and Eight are other computer optimised, narrow-band designs. Gain figure-of-merit is antenna gain minus maximum practical gain (maximum gain for the boom-length with reasonable input impedance and bandwidth, a definition which is vague but which can be evaluated mathematically as a function of boom length). As N6ND says, for a really good signal you need dBs in the air and dBs on the desk. This simple Yagi design will take care of airborne dBs. Desktop dBs are a matter between you, your licensing authority, and your spirit of adventure.

Element Postitions and half lengths for Optimised 5-element Yagi (inches)

Position	E1-#1	E1-#2	E1-#3
0.00	57.54	57.42	58.43
44.06	54.85	54.59	55.39
107.89	53.48	53.16	53.88
194.86	52.59	52.23	52.89
270.62	52.59	52.24	52.90

Element Positions and Half-Lengths For Optimized 5-Element Yagi

El #1: 0.375" diameter, non-conductive mounting bracket

EI #2: 0.5" diameter, non-conductive mounting bracket.

El #3: Cushcraft A50-6S element mount 24" half-length of 0.75" diam., 0.625" diam. tips, U-bolt mount

Half-lengths are lengths measured from the centre of the boom to the element tip.

To return to the archive page click here

SIX METERS

ARRL ANTENNA Vol 5 COMPENDIUM contains an article I wrote providing details about two portable 6 meters antennas I developed, a two-element quad and a three-element yagi with telescoping elements.

Both antennas were designed for easy construction and quick assembly and disassembly. The quad provides a measured gain of 4.2 dB over a dipole and the yagi 5.8 dB over a dipole. Details are given describing the methods used to measure the gain of both antennas.

Here is a photo of the 2 element quad taken when camping by the ocean. I am looking for new grids.

The following picture was taken on a beautiful summer evening while operating portable. The 6 meter portable yagi with collapable elements is clearly visible against the evening sky.

I have scanned a copy of the article "Two Portable 6-Meter Antennas" here Reprinted with the persmission of the ARRL. Copywrite ARRL. All rights reserved.

ARRL ANTENNA Vol 5 COMPENDIUM is still available (January 2003) from ARRL (208 pages. First edition, © 1996-99, The American Radio Relay League, Inc. (ISBN: 0-87259-562-5) #5625 -- $20.00 USD:

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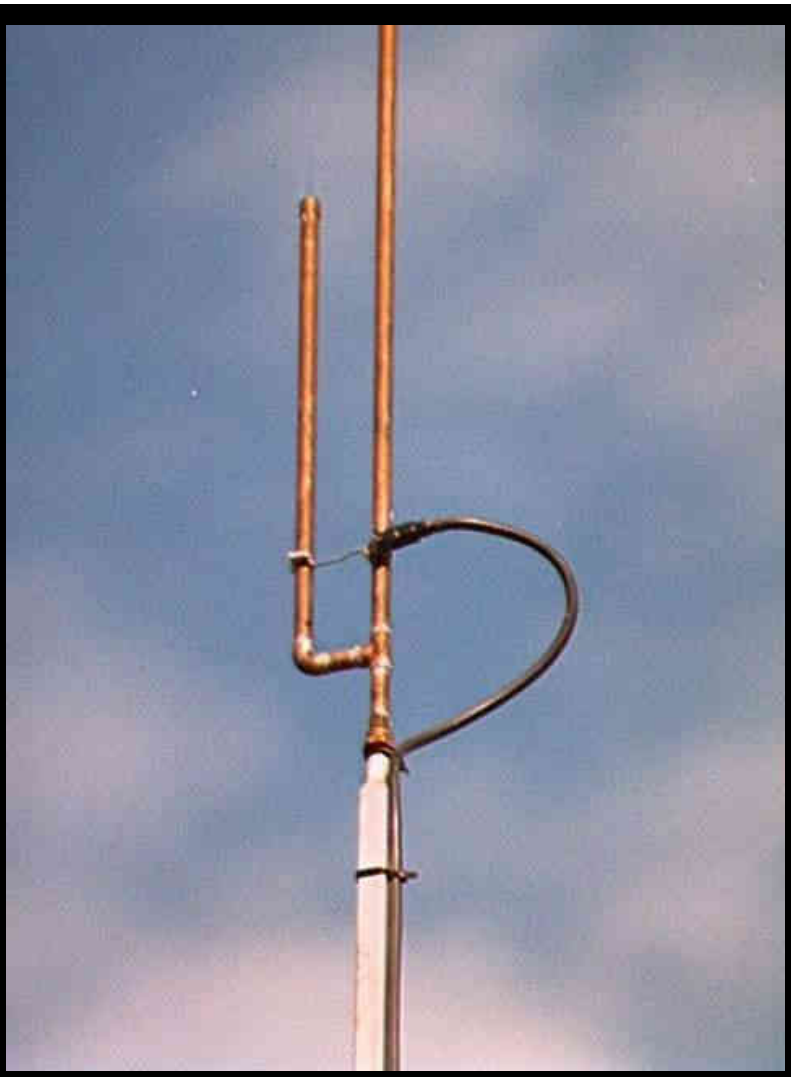
Copper Cactus Dual-Band Super J-Pole Antenna Project

Your going to like this super J-pole project, This home made j-pole is easy to build and sturdy, While looking for an jpole antenna project to build I remembered seeing a Marine antenna called the Super J-Pole in the 1988 ARRL Antenna Handbook, which claimed a 6 db gain over a quarter wave ground plane. I didn't have a machine shop at my disposal to fabricate the parts shown in the Marine antenna article so I set about redesigning the antenna using materials that were easy to find and work with. I have had very good results working with copper J-Poles, so I built my refined version of the classic J-Pole. I then added a short insulated section, the extra half wave of vertical length, and the needed half-wave matching stub.

Here is a schematic drawing of the home brew J-pole (Click Here) Word doc

Materials

Put the Heat to it



All the materials except the S0239 fitting can be found at any good hardware store, and the whole antenna can be made in less than an hour.

In my design I use 1/2" copper schedule M tubing and 1/4" soft copper tubing. I had experimented with using Teflon insulator, but have since changed my design to use a 9" length of hardwood dowel with three coats of lacquer as the insulator, for more strength.

Clean all the tubing, and then from the 1/2" tubing cut one piece each of the following lengths: 57-1/2"; 38"; 19"; 2", and a piece about 3" long for a stub to mount the antenna. In addition to the tubing, buy a 1/2" elbow, a 1/2" Tee, two 1/2" end caps a 1/2" threaded fitting, and a cast iron floor flange for mounting. Get a piece of 3/16" or 1/4" soft copper

Now fire up the torch and start the assembly process from the bottom. See Figure 1. Use flux on all joints, solder the 1/2" threaded fitting to the mounting stub, and solder the 1/2" Tee fitting. Then proceed with the 57-1/2" section, 2" cross piece, and 19" section. Pay close attention to getting the 19" piece parallel to the 57-1/2" piece. After these have cooled, drill through both the 57-1/2" section of the 1/2" tubing and the hardwood dowel about 1/4" from the top end of the 1/2" tubing, and the bottom of the 38" section of the tubing. (See Photo A.) Then insert the 1/4" tubing to the 1/2" tubing and sweat solder the end caps. After these have cooled, clean the entire antenna, bend the half wave matching section to a half circle of about 4" radius around the antenna to help the balance and match.

tubing 42" long. Find the center of the 1/4" tubing and bend it around a 1"-to-1-1/4" diameter water pipe or dowel.

Simplify the Feed Point of the J-pole

The feed point also needed to be made simpler, so I elongated one of the mounting holes of a panel mount SO-239 fitting and inserted a stainless steel adjustable band clamp. This goes on the 57 1/2" long section of 1/2" tubing. A short 2-3/4" length of # 14 copper stranded wire is soldered to the center terminal to go over to the 19" section. I used another stainless clamp to attach this. (See Photo B.) While experimenting to find the proper feed point, I found that the distance above the crossbar should be about 3".

Part's List

- 1 10-foot section of schedule M 1/2" copper tubing
- 1 1/2" copper elbow
- 1/2" copper Tee fitting
- 2 1/2" copper end caps
- 1/2" copper threaded fitting (for mounting)

Building Suggestions

1. You may use a Fiberglas rod as an insulator, but you will have to be very careful with the torch or you may weaken or burn the rod, or make it brittle.

2. When cutting the 1/2" copper tubing, cut the 57-1/2" piece from one end of the 10" length, and the 38" piece from the other end. By doing this you will have factory-cut edges for inserting the 1/2" dowel.

3. Be sure to keep the flame of the torch away from the insulator to avoid burning it.

4. Use paste flux on all joints when fitting the pieces together. Use enough flux, since you will be cleaning the entire antenna with solvent after assembly.

5. Use a weight to hold the 19", 57-1/2", and 2" pieces, and the Tee and the elbow, flat when they are sweat soldered together.

6. Use a ruler or caliper to check the spacing between the 19" and 57-1/2" pieces, to keep them parallel to each other.

7. When drilling the SO-239 fitting, use the drill press. Be careful not to drill into the threads of the fitting. After the holes are drilled, file the opening flat for a better band

- 1 1/2" cast floor flange (for mounting)
- 1 Piece of 3/16" or 1/4" soft copper tubing 42" long
- 1 Piece of 1/2" hardwood dowel of Fiberglas rod
- 1 SO-239 panel mount coaxial fitting
- 1 Piece of # 14 stranded copper wire
- 2 3/8" by 7/8" stainless band clamps

Tools needed:

- Tape measure
- Tubing cutter
- Propane torch
- Solder and flux
- Electrical tape
- Caulking compound
- Screwdriver
- A weight to keep parts aligned while soldering
- Steel wool or a Scotch Brite pad (for cleaning all copper)
- Spray can of clear exterior lacquer (to finish-coat completedantenna)

clamp fit.

8. After the best match has been found, you may want to solder the SO-239 and the stranded wire end to the 1/2" tubing.

9. When the antenna has been cleaned and matched, spray the entire antenna with a coat or two of clear lacquer to keep it looking nice.

10. After everything else has been done, apply silicon or a butyl rubber compound to the insulating section, then cover the joint with electrician's tape for a weather tight seal.

11. A 1/2" pipe coupling and a length of pipe may be used in place of floor flange for mounting in a roof tripod.

Suggestion by KB9TIO

An Improved(?) feed arrangement is to use copper couplings, same size as the tubing. Split the coupling lengthwise, bend out @ 1/4" and notch for the so-239 inner ring. attach with sheet metal screws an/or solder so-239 to coupling. repeat with 2nd coupling and solder other end of #14 wire to flat on coupling. this provides a strong joint and very easy tuning by sliding the couplings up or down the tubing as needed. this tip was given to me by Tom-KB9OZZ

Thanks, Matt KB9TIO

Suggestion by **KD7GQC**

I have also seen variations on the J Pole which omit the insulator section. In the first, the center conductor is soldered directly to the driver element, and the entire antenna is thus at DC ground. The second variation is also an all-copper design, but you capacitively couple the RF to the element, by wrapping several turns of insulated wire from the center conductor around the driver section of the J pole. I can't vouch for the second approach, but the first seems to work great, for installations where a DC "short to earth" is not a problem. It seems to me that eliminating the insulator section makes for a simpler, and potentially stronger, design. Kudos on a great antenna! 73s, Brian KD7GQC

One last suggestion: put a piece of steel about 1/2 or 3/8 of about a foot and a half up the center of it (from the base) this will strength the bottom portion of the pole.. put a small curve in it so it touches the inner wall of the pipe.. my parents use this j-pole and live in a windy area.. and this fixed that problem.

In order for these instructions to work great.. Please send me feedback on if you had problems and what can be done to help the next ham.. I don't get much feedback on the copper cactus J-pole, Guess that's good in a way, but would love to either way **Praise or Suggestions** about this project.

Recommend This J-Pole to a Friend

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QHtenna by N4QH

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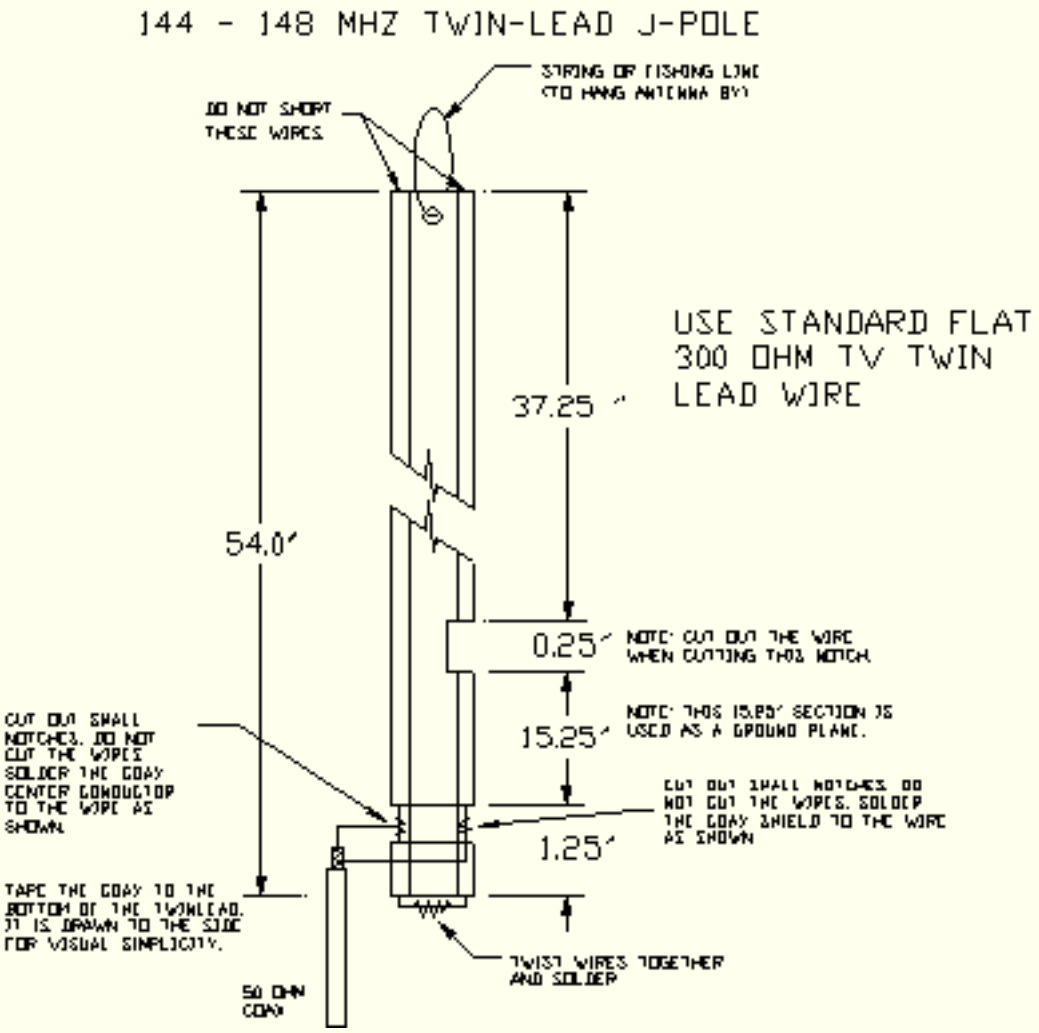
Santa Clara County, California
ARES/RACES

Emergency Antennas

- ***300 TV Twin-lead J-Pole Antenna (Approximately 3 dB Gain)***

  This is an easy antenna to make from existing or inexpensive materials. It rolls up for easy storage and can be deployed in seconds; just hang from the nearest ceiling or attach to the end of a fiberglass fishing pole or PVC pipe. Better than a rubber duck (what isn't?).



- ***¼ wave Mag Mount as a Portable Antenna***

If you need to operate portable and only have a mag mount antenna available, try placing any large piece of metal underneath it. This might be a refrigerator, stationary car, or a metal rain gutter. You might also try making your own portable ground plane by placing some aluminum foil over a large piece of cardboard (2 feet x 2 feet minimum).

- ***American Legion J-Pole***

  If you are able to operate fixed portable, try the "American Legion J-Pole". Attach it to the top of a 10 foot section of PVC pipe and mount this to a camera tripod. Attach weights to the legs for stability. This makes a very nice fixed station antenna.

To add your favorite antenna to the list, contact the Webmaster

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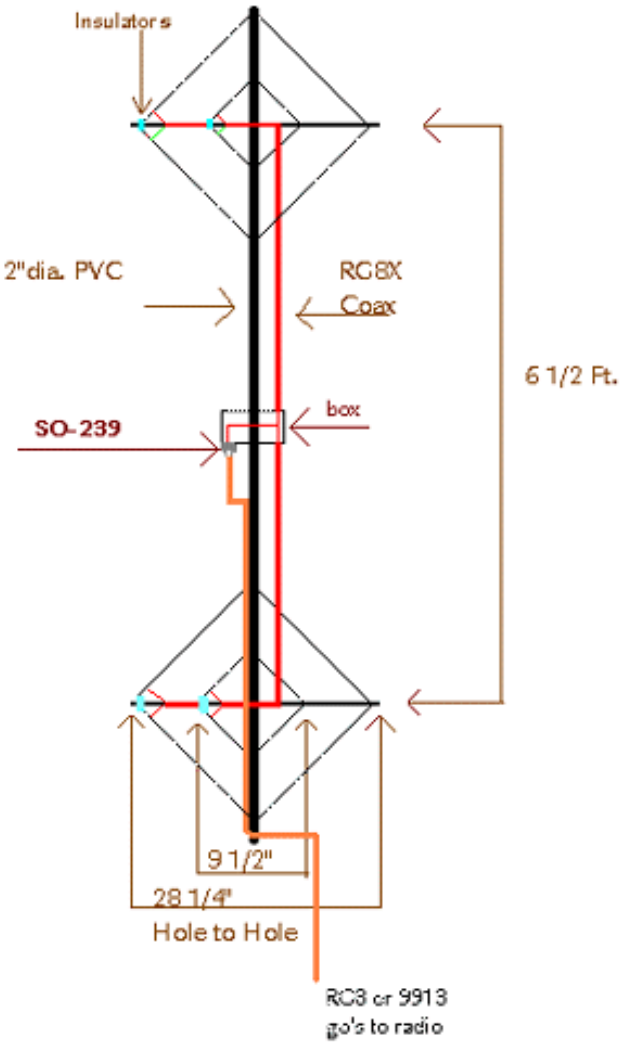
Web Site Home Page

This page was last updated04/20/02

Specifications for antenna at bottom of diagram



KE4UYP's Biquad

144/440 MHZ

Specifications

This is a vertically polarized Broadside and Collinear Array each Loop is a vertically polarized Broadside Array both the right and the left side of the Loop is a bent 1/2 wave dipole working 180 degrees out of phase with the other one and that gives you 4.15dbi of gain The two Loops are Stacked one above the other this makes a Collinear Array at one wavelength at 2m that gives you 5.15dbi of gain so the Total gain of the Antenna is 9.15dbi all this from four pieces of wire and some PVC pipe.

One of the easiest construction techniques for building this antenna is to use 3/8" fiberglass rods for the horizontal spreaders that pass straight through the two inch PVC pipe you can hold them in place with two nylon tie wraps one on each side of the two inch PVC pipe pulled tight around the

fiberglass rod.

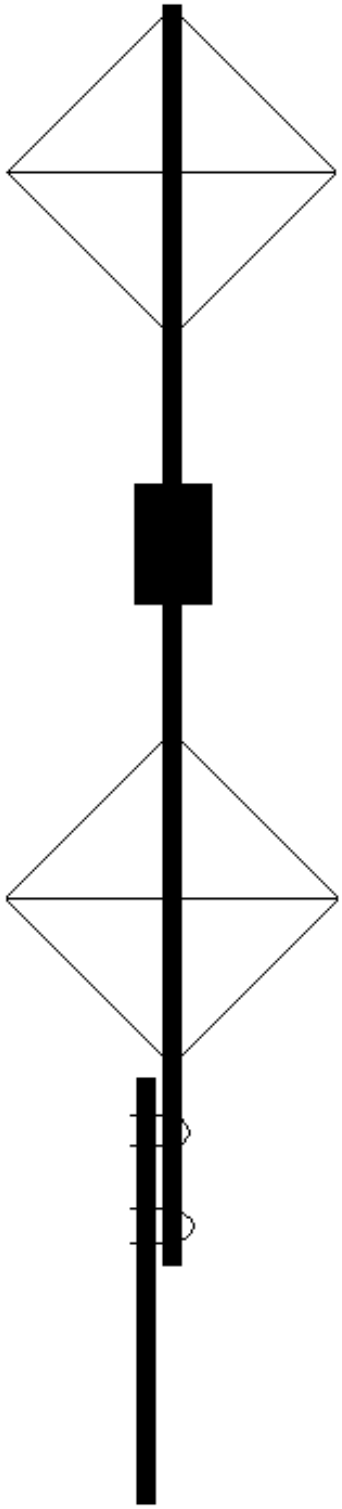
You can make insulators from half-inch diameter PVC pipe 3 inches long. Drill a 3/8" hole 11/2" from one end of the pipe all the way through both sides of the pipe, then drill two 1/8" holes near the ends of the PVC pipe for the antenna element wire to go through. Then you can slide this pipe on to the fiberglass rod.

If you solder the coax connections then seal it with epoxy tape it will be 100 percent reliable and weather proof, also make sure you extended the two inch pipe twelve inches past one of the loops you can U bolt that part of the pipe to your mast, of course this would be for FM operation. See attached drawling.

This is a inherently broadbanded antenna so fine-tuning the elements is not necessary the reason why these elements are shorter than a standard element is because when you connect two driven elements in parallel with each other this creates and an excessive amount of inductive reactance and the easiest way to remove this is to simply make each element shorter normally these elements would be three inches longer on 2 meters.

The impedance of one loop is 100 ohms so when you put the two loops in parallel you get 50 ohms this makes a perfect match for coax cable. If you are using this for 2 meter SSB then of course you need horizontal polarity and the loops would be side-by-side this puts the feed points at the bottom corners of each loop. the antenna would then be mounted on the mast from the center.

The formula for calculating the overall loop lengths is 11665.4 divided by xxx.xx Mhz's the answer will be in inches.



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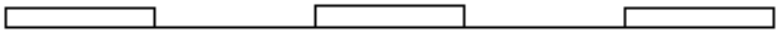
The Simple Collinear

Wherever there is a need for a vertically polarized omni-directional antenna with gain, a collinear is used. The vertically polarized collinear takes the form of two or more half waves in-line and in-phase, resulting in an omni-directional pattern with maximum radiation near the horizon. The techniques used to ensure that the half waves are in phase are:

1. Interspersing the in-phase half waves with quarter wave stubs. This is the method described in the ARRL Antenna Book.
2. Making the antenna out of half wavelengths of coax with the outside of the coax used as the radiating elements and the inside of the coax used as the phasing elements.
3. Interspersing the half waves with resonators to provide the necessary phase reversal.

Each of these techniques is designed to reduce or prevent radiation from the out-of-phase components of the antenna. The Simple Collinear uses a different technique.

The Simple Collinear makes use of the fact that large diameter elements within an antenna can radiate more readily than small diameter elements. Therefore if the large diameter elements are in phase and the small diameter elements are in phase but out of phase with the large diameter elements, radiation from the large diameter elements dominate.



Shown above is a schematic of a 5 element collinear—3 large diameter elements and two small diameter elements. Each element is about ½ wavelength long, and since the phase switches every half wavelength, the 3 large diameter elements are in phase and the 2 small diameter elements are in the opposite phase. The result is an antenna with about 2 dB less gain than if all the elements are in phase. Simulations show that this can be applied to very large arrays. A 45 element array (23 large diameter and 22 small diameter elements) is simulated to have 14.38 dBi gain.

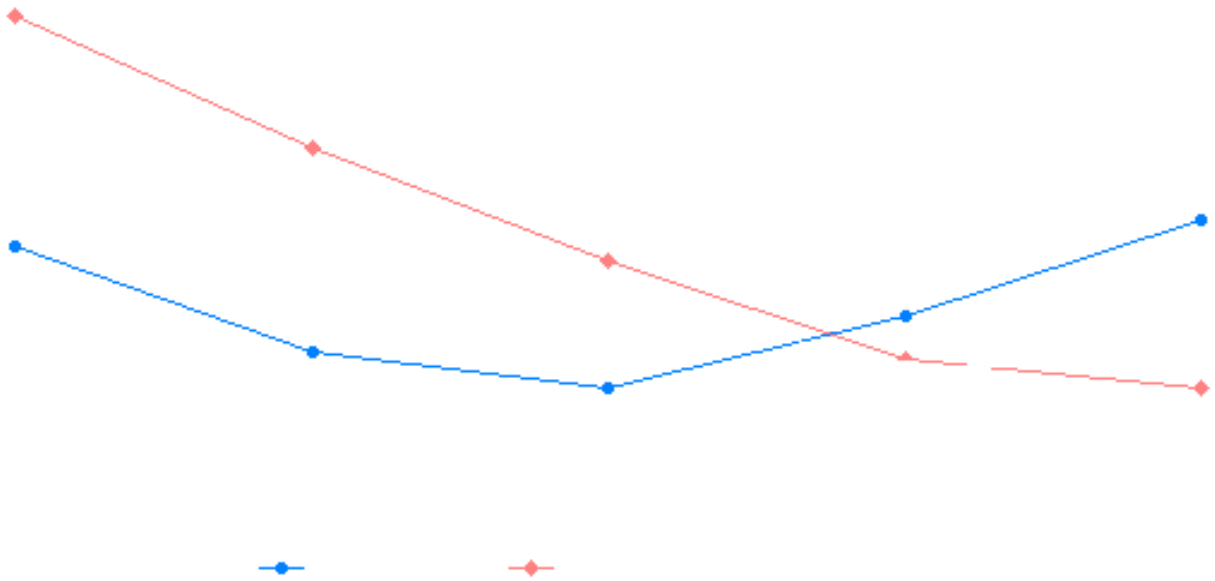
Performance vs Number of Elements			
# of Elements	Gain (dBi)	Z (ohms)	Ideal Gain
1	2.06	79 + j14	2.06
3	3.76	103 + j7	5.41
5	5.17	117 + j1	7.32
11	8.09	144 - j14	10.49
23	11.33	176 - j42	13.6
45	14.38	207 - j90	16.48

Large diameter elements = 1 inch.
Length of large diameter elements = 38.15 inches.
Small diameter elements = 0.040 inch.
Length of small diameter elements = 40.95 inches.

Ideal Gain assumes all elements have
1 inch diameter and are fed in the center.

The table above shows the gain and impedance of a 2 meter 45 element antenna, consisting of 23 elements, 1 inch diameter and 38.15 inches long and 22 elements 0.040 inch diameter and 40.95 inches long. The antenna has its feed point the center of the lowest element, although it could be fed at the center of any of the 45 elements. Also shown in is the performance of various versions of the same antenna, varying the number of elements between 1 and 45. Note that the gain increases by 3 dB for each doubling of the number of elements. I stopped the simulation at 45 elements because I can't imagine building an antenna larger than this—at least at this time. Also shown in this table is the simulated gain of an ideal antenna assuming all 45 elements have 1 inch diameters and are fed in phase with voltage sources. Note that gain is only about 2 dB higher than the simulated real antenna. This is because the in phase currents are distributed quite uniformly across the antenna, an effect achieved by adjusting the lengths of the elements slightly from half wavelength. The 0.040 diameter is a compromise between making the diameter small to reduce the radiation, but not so small that ohmic losses become important. The 45 element array is simulated to have only 0.1 dB loss due to ohmic loss in the 0.040 diameter copper elements. All these values were determined using K6STI's AO Antenna Optimizer MININEC program.

A 5 element test antenna was simulated and built. It was optimized for 5 elements and not created from a reduction of the 45 element array in the table; its gain was simulated to be 5.7 dBi and impedance about 100 ohms at 147 MHz. The small diameter elements were simulated to be 0.015 inch because for a small array, radiation from these elements is more of a factor and ohmic loss is less of a factor. All elements were 38 inches long. The 1 inch elements were made from aluminum, but any metal will have negligible loss—copper, aluminum, galvanized, or steel are all ok. The small diameter elements should be 0.010 to 0.020 inch in diameter (#30 to #24). There will be a slight reduction of gain if #24 wire is used. I used #26. Support of the antenna in the small diameter sections was provided by ¾ inch diameter schedule 40 PVC pipe. Guying was provided by 3/16 inch polyester cord. The antenna was fed through a quarter wave transformer made from RG59U. A few ferrite beads on the RG59U provide any needed current balun action. The antenna was installed on the roof in place of an old TV antenna. Simulated and measured SWR, with measurements taken at the end of the transformer, are shown.



This antenna has many attributes. It is easy to model. It is easy and inexpensive to build. A high gain version can be built in-situ, adding a couple of elements at a time, adding guying as elements are added. It is broadband. Even the 45 element version is simulated to have an SWR less than 1.7 across the 2 meter band (144-148 MHz). Current distributes well over the entire antenna making the aperture efficiency high.

Here are some possible applications.

1. Small versions like the 5 element antenna described can be roof mounted for repeater access.
2. Versions of this antenna should be good for applications in the field such as for contesting or DXpeditions, giving good gain and good height cheaply.
3. This antenna would make a good base station antenna for 2 meter and higher repeaters. Down tilt can be designed in if desired by slightly shortening the elements.
4. It should be a good antenna for communicating via Tropospheric propagation.
5. Several could be put together to make a phased array.
6. It could be used as a feed for a cylindrical parabolic antenna. Large cylindrical parabolic antennas should be easier to make than large paraboloid (parabola of revolution) antennas.

Ross Anderson W1HBQ September 7, 2002

E-mail me at ross_anderson@comcast.net

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K6STI can be reached at bb at n2 dot net.

My homepage “Ross’s Antennas” with links to my other pages. http://home.comcast.net/~ross_anderson

Keywords: The Simplest Collinear

Build a 2 Meter DDRR for Mobile

By W5GVE

This is a story of an antenna and its modification over time into "better" forms. Part of the improvement was by design, and part was just plain good luck. When you are working closely with an antenna, unless you keep very detailed records, it is difficult to discern which is which after the fact.

PURPOSE

The original objective was to build an antenna for 2 meters FM (146-148 MHz), vertically polarized, and omni-directional. The world is already full of such antennas; why something new? The truth is that I needed a short antenna that wouldn't scrape against the low-hanging trees of the desert where shade is precious, or against the 7 1/2 foot ceiling of the multi-story parking garages in the cities I occasionally visit.

Many years ago, I was intrigued by the DDRR antenna. It originally was a quarter wavelength radiator sprouting up from a ground plane, taking a sharp turn to the horizontal and coming around in a circle to almost meet itself where it sprouted. Sort of an open loop parallel to the ground with one end connected to the ground. This obviously horizontal antenna behaved in an outrageous fashion—***it radiated vertically!*** There's no free lunch however and, as originally described, it worked about half as well as a full-sized quarter wave vertical wire. But, if you could stand the tariff, it ***was*** low to the ground—$1/10^{th}$ as tall as the vertical.

I stood it as long as I could and finally built one for the Citizen's Band (27.5 MHz). This was years ago, and as I recall, the mobile CB-DDRR was three feet in diameter, six inches tall and looked like a pair of hula hoops. It was built of 1/2" aluminum cable-TV trunk cable held together at crucial points with automotive hose clamps. And, it worked! At least ***until my wife saw it,*** and said she could not be seen dead driving around town under it! So, DDRRing went dormant for a decade.

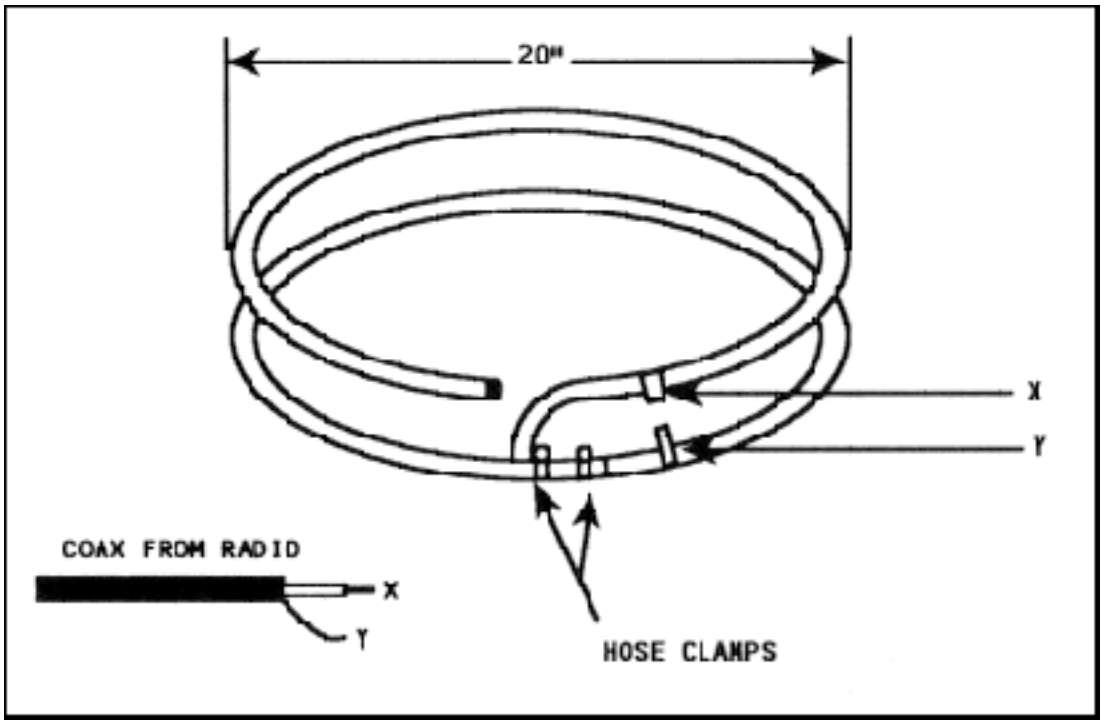
In subsequent years, more articles appeared on the DDRR in its quarter-wave version, and later, a half-wave model. I didn't try either and pretty well believed what I was reading. As the need for the 2M low profile antenna grew, my mind turned to the 1/4 wave DDRR, throught the 1/2 wave DDRR and ended on the great idea of maybe a 3/4 wave DDRR!

A 3/4 wave edition would not be any taller than a plain vanilla (1/4 wave) DDRR, and would have much greater capture area. It seemmed like the perfect solution. If it looks weird, so what? My wife has her

own car now. All systems were GO, so I began. Not having a whole lot of imagination, I began by resurrecting the concept of the CB-DDRR, 1/2" aluminum TV cable and all.

At 147 MHz, a wavelength is quite close to 80" and 3/4 wavelength is 60". The diameter of a 3/4 wavelength DDRR is 20". I arbitrarily decided that 3" in height (center to center) would probably be tall enough to help capture incoming wave fronts, and it ***did*** resemble the old CB version.

Figure 1



WORKED LIKE WOW!

For testing purposes, I taped it to the roof of my little Chevrolet station wagon. Wow, did it ever work! I compared it to the 5/8 wave vertical magnetic mount whip on the other end of the station wagon roof, and the 3/4 wave DDRR equaled or exceeded the commercial whip. One little problem. It had a great gain off one section of the antenna, average elsewhere, and a tiny area of a slight null. Further tests showed it had a substantial amount of horizontally polarized radiation in addition to the desired vertical radiation. What was going on? The Mark I version has problems!

DIRECTIVITY

I believe the 3/4 wave DDRR showed directional characteristics because its physical structure is about 1/4 wavelength in diameter. Current on exactly opposite points of the radiator's circumference are 3/8 wave (30") out of phase, electrically, but are 1/4 wave apart (20") physically. I did not have the means to work this out theoretically, but it seems reasonable that this physical configuration would simulate radiation patterns of discrete radiators phased and separated in this manner. I decided this was the kind of a problem that could not be overcome. It could be developed if, some day, I want to build a directional vertical low-height radiator. Maybe some day.

POLARIZATION

The method of feeding the antenna was the same as I used years ago on the CB-DDRR. Since the DDRR can be looked upon as a section of parallel conductor feedline, with one side joined to a ground plane, one end is effectively shorted and presumably at a very low impedance point, and the other end is at a very high impedance, it seemed reasonable that at some point near the low end there are a pair of points at 50 ohm—just right for direct coax feed. It worked on the CB-DDRR and I figured it would work on the 2M version too.

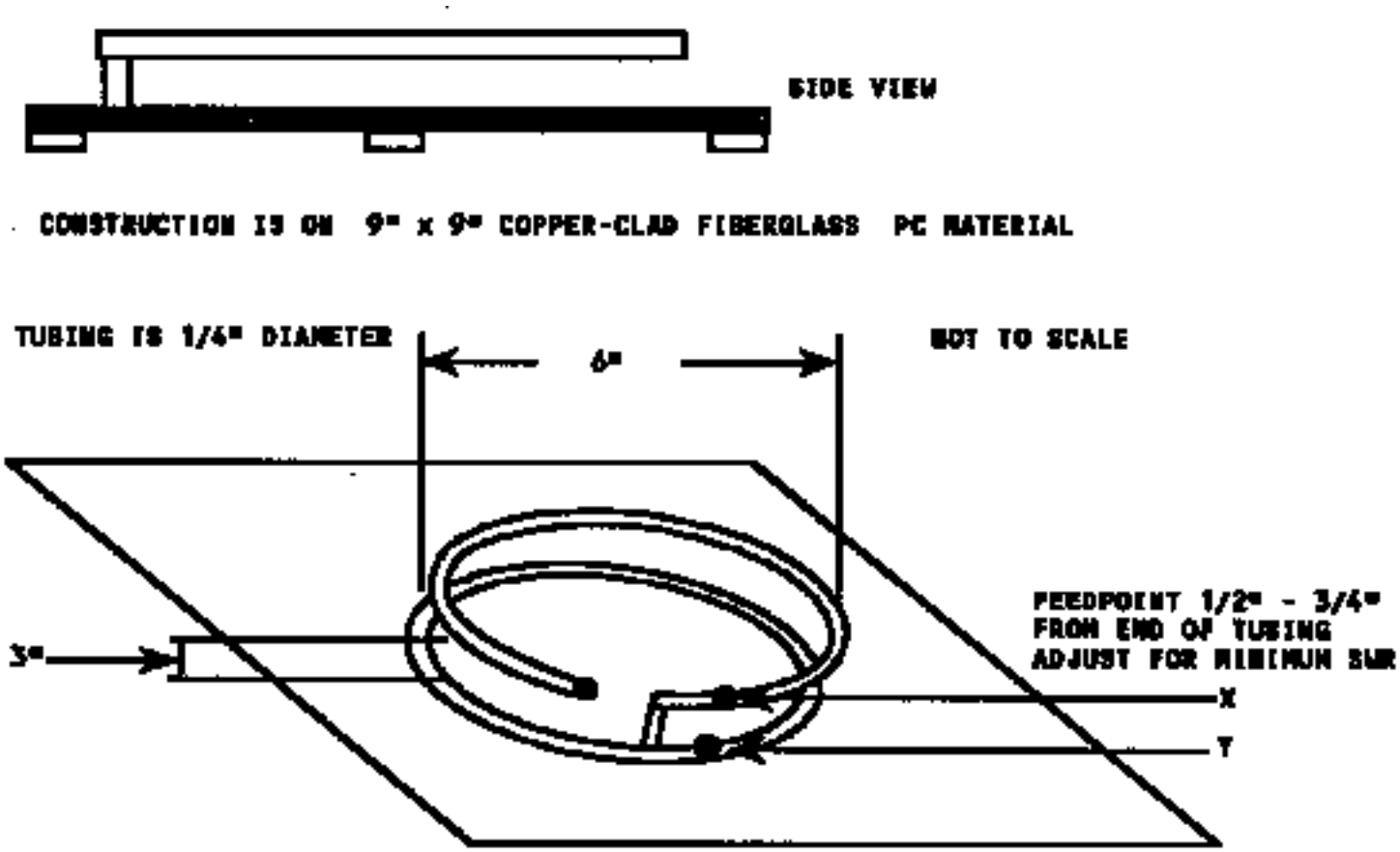
A straight transmission line antenna, laid out horizontally, would radiate horizontally polarized energy. But when the same antenna is curled up into a loop, about 1/2 wavelength in diameter, it radiates vertically polarized energy. In order to get a good match (50 ohms, with low reflected power) I had to skew the feed points slightly. That is, the center conductor of the coax was connected to the upper line at a point NOT directly over the point where the coax braid is connected.

TOO MUCH SKEW?

I believe the skewing of the feed points resulted in the mostly vertical, but partly horizontally polarized radiation. Again, this is an area for further eXperimentation, since I did not pursue it at this time. So much for the 3/4 wavelength DDRR.

Considering the phasing problem that gave me the unwanted directivity, I concluded that some of this could be expected with a 1/2 wavelength. So, I decided to do a plain old 1/4 wavelength DDRR, but by using some of the information about transmission line radiators I had picked up in the intervening years since the CB-DDRR, mainly the work of ***Ted Hart, W5QJR.*** The better the conductivity of the radiator, the better the antenna. That rules out aluminum tubing and hose clamps.

Figure 2



SOME SUCCESS

I decided to begin with a 1/4" diameter copper tubing, formed into two 6" diameter loops, separated 3". One loop is tack-soldered at 3" intervals to a 9" square of copper-clad fiberglass printed circuit board stock. The coax braid is soldered with no length to the circuit board. The center conductor is taken up 3" and soldered to the free loop about 2" from the shorted end. Matching is accomplished by three methods: adjustments of the tap on the free loop; trimming the free end of the loop (raising frequency of resonance); and bending the free end of the loop up or down (raising and lowering, respectively, the frequency of resonance). The results? Equal to the 5/8 mag-mounted vertical whip! The Mark II is a success. Now to improve on this!

THE NEXT VERSION

The Mark III version was also begun on a 9" square of circuit board material. It has great surface conductivity, is stiffer than solid sheet copper, can be soldered readily, and can be attached to a magnet for car-roof mounting. Instead of a double loop of 1/4" copper tubing, the Mark III uses a single 20" long loop of 5/8" copper tubing, a 5/8" copper "L", and a 2" piece of 5/8" tubing to space the loop over the ground plane. Matching is the same as the Mark II's, with the exception of soldering to the loop. I had to use a fancy homemade copper clamp-low resistance, moveable, and IT CAB be soldered!

Once the best tap point has been determined, drill a small hole at the point. Use a sheet metal screw and a copper tab to solder on the coax center connector. The adjustments for best match are limited to two: trimming the end of the loop with a tubing cutter and moving the feed tap. This is a high-Q device, and moving the feed tap has a very large influence on the resonant frequency, as well as feed point

impedance. Performance? Equals or exceeds the 5/8 wave vertical whip in all respects. Being fairly high Q, it has a useful bandwidth, a little narrower than the whip—about 2 1/2 MHz.

Figure 3

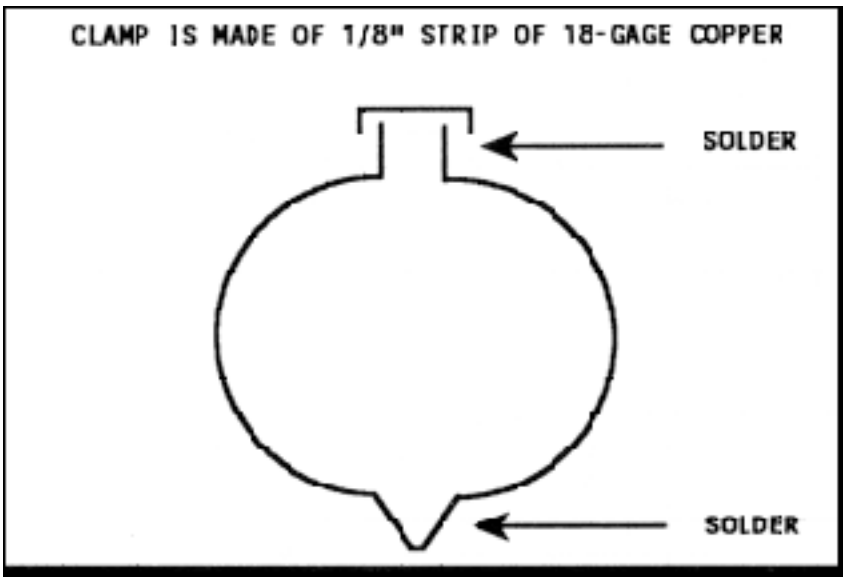
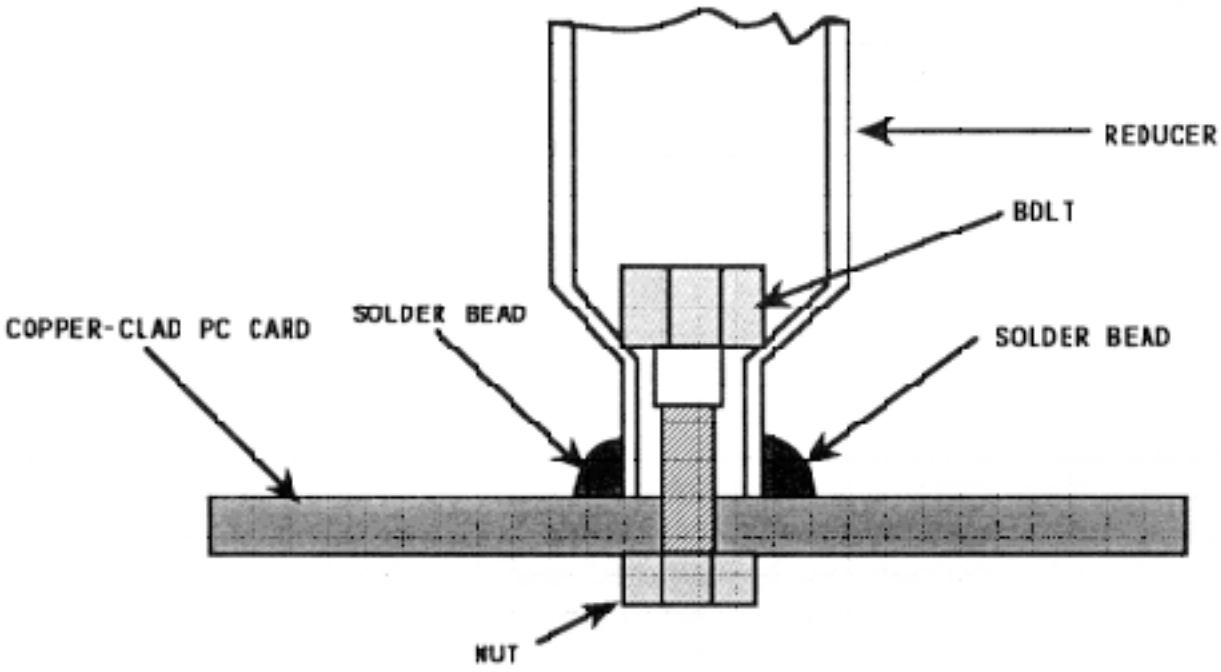
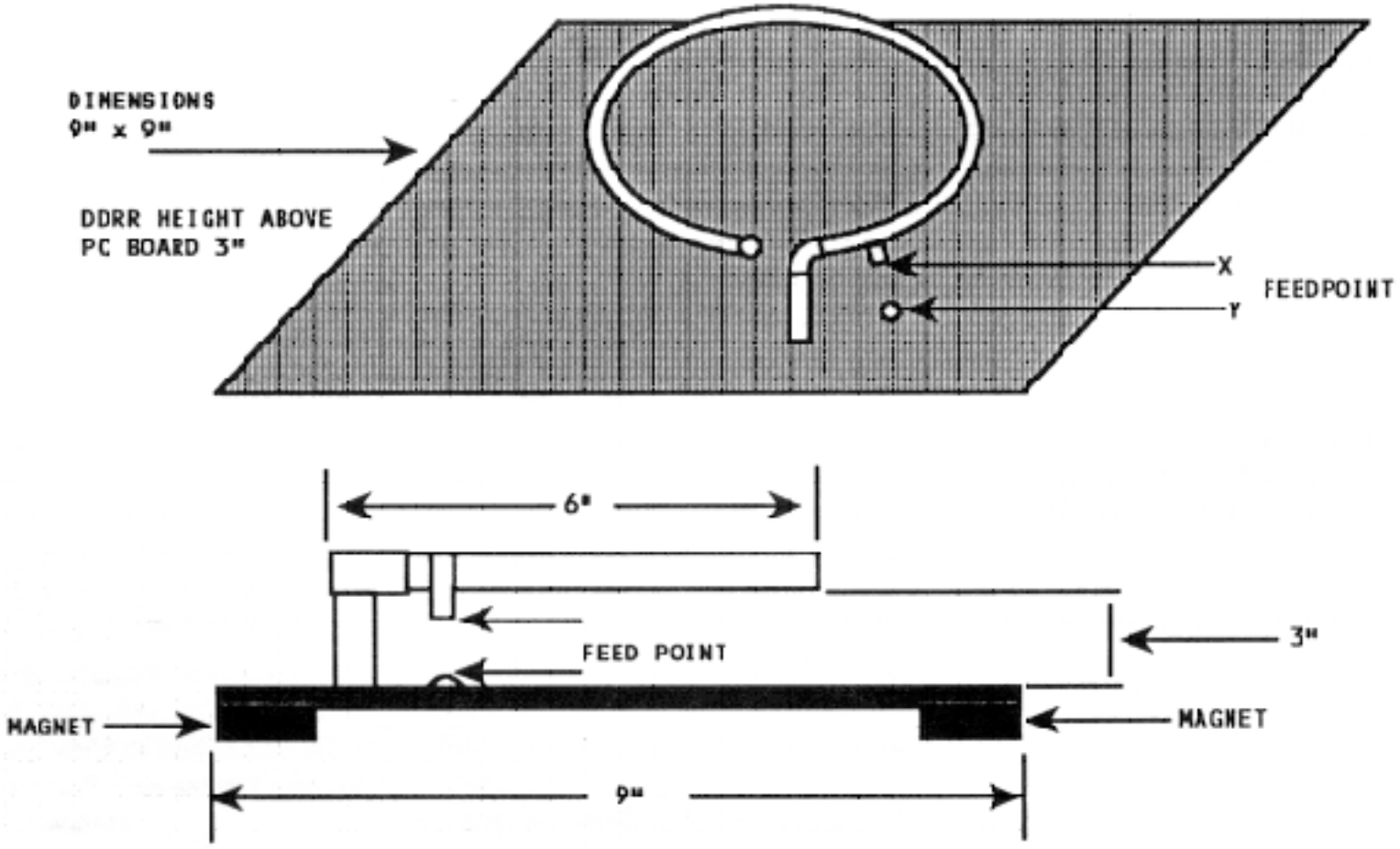


Figure 4





SPECIAL CONSTRUCTION DETAIL

Fixing the butt of the free loop of the Mark III to the ground plane took some real effort. I finally solved the physical problems by clamping a bolt inside the 2" vertical piece of 5/8" tubing in the jaws of a vise and clamping the tubing again below the bolt head to form a shoulder against which the bolt can be pulled tightly, securing the loop to the ground plane. Gently solder the butt to the circuit board, flaming the tubing and not the PC board. **-30-**

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Last modified: July 04, 2004

The Cycloid Collinear, a CP Omni

Figure 1

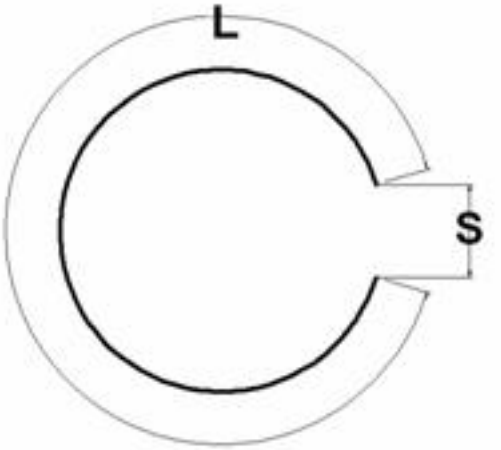
Despite the advantages of circular polarization, amateurs have made little use of it. One of the main reasons is the difficulty in making a good circularly polarized omni-directional antenna (cp omni). This paper describes a wide band, high gain, multi-element cp omni with a single 50 ohm feedpoint with instructions for building a 2½ element cycloid collinear for 2 meters.

There are no commercially available cp omnis for amateur use [1], although there are many available for the fm broadcast band [2,3]. A common design is the cycloid which consists of a horizontal loop approximately lamba/2 in circumference, with short vertical extensions at the loop ends [2,4]. The radiation from the horizontal loops is approximately omni-directional and approximately equal to the radiation from the vertical sections (and is 90 degrees out of phase), resulting in circular polarization. The impedance of this antenna is not 50 ohms and requires a matching section [4]. If more gain is desired, then two or more elements can be stacked collinearly, with each element fed separately [2].

A simpler way to combine elements is to extend the vertical sections of each element until they touch. However doing this increases the vertical radiation and the antenna becomes mostly vertically polarized. Simulations have shown that this makes a very good vertically polarized omni with a small amount of horizontal polarization. The vertical polarization radiation can be reduced by reducing the diameter of the vertical wires [5]. The antenna described here is just such an antenna.

A test antenna was modeled for 146 MHz consisting of three horizontal loops

made of 3/8 inch diameter soft copper tubing and two #18 (0.040 inch diameter) vertical wires. Mininec is required for the modeling because nec cannot simulate an antenna made of two wires of very different diameters [6,7]. The design goal was to design a right hand circularly polarized antenna with 50 ohms impedance at 146 MHz.



A loop drawing is shown in Figure 2. For the top and bottom loops, L = 33 9/16 and S = 1 ¾. For the center loop, L = 28 and S = 1 ½. Cut the copper tubing to the lengths shown and bend them to approximate a circle with a gap as shown. All dimensions are in inches.

Then support the elements in some way. I used a ten foot length of ¾ inch schedule 40 pvc pipe. The spacing between the elements (and the length of the vertical wires) should be 57 7/8 inches. Figure 3 shows how the vertical wires and the loops are connected together. This figure is drawn from the point of view slightly off axis from the top or bottom of the antenna. The polarization from an antenna assembled in this way is right hand circular polarization (rhcp). The antenna is fed in the center of the center loop through a balun.

Figure 1 shows the assembled antenna on the roof. No patterns were taken, but the good agreement between the simulated and measured SWR (Figure 4) gives confidence that the antenna is working properly.

The simulated gains at 146 MHz are:

Right hand circularly polarized gain = 4.54+/-0.59 dB.

Vertically polarized gain = 2.76+/-0.05 dB.

Horizontally polarized gain = 0.63+/-1.45dB.

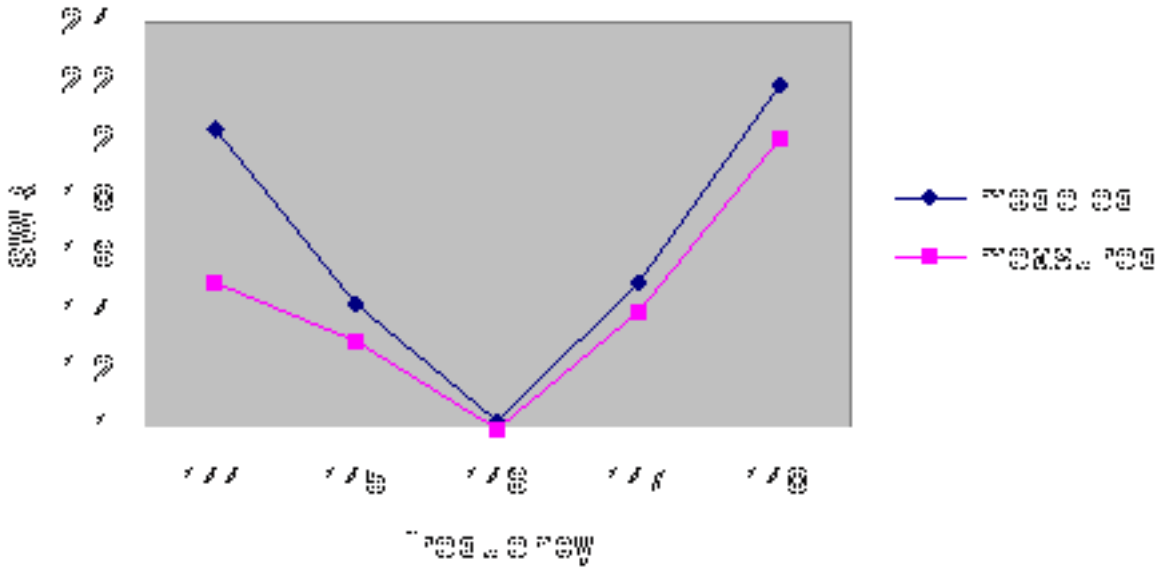


Figure 4

Adding more elements will improve the gain. If you decide to change the design goals using different diameters, more elements, or different frequency, you should resimulate using mininec. If you decide to make a mostly vertically polarized omni using the same diameter for the vertical and horizontal elements, then nec (which is more accurate than mininec) can be used.
I have not said anything about why you would want a cp omni. Please refer to the references for discussions on circular polarization.

Ross Anderson W1HBQ April 19, 2004 ross_anderson@comcast.net

References and Links

[1] GBPPR Tech Bulletin #2 - Advantages of Circular Polarized FM Repeater Antennas http://www.qsl.net/n9zia/cir_pol_rpt.html

[2] http://www.progressive-concepts.com/37a.html

[3] http://www.psibroadcast.com/antenna-fm.asp

[4] The WA7X Beacon The Cycloid Dipole
http://www.wa7x.com/cycloid_info.html

[5] The Simple Collinear http://home.comcast.net/~ross_anderson/sc.htm

[6] When MININEC is Superior to NEC
http://www.cebik.com/amod/amod56.html

[7] I use K6STI's AO. His e-mail address is bb at n2 dot net.

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http://home.comcast.net/~ross_anderson/quadix.htm

My homepage “Ross’s Antennas” with links to my other pages is http://home.comcast.net/~ross_anderson

keyword: omnidirectional

The HO Collinear, a Horizontal Omni

Figure 1

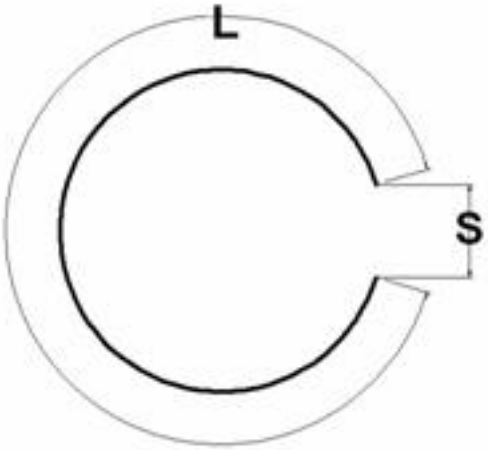
This paper describes a high gain, multi-element horizontal omni with a single 50 ohm feedpoint with instructions for building a 4 element array for 2 meters.

There are several ways to make a horizontal omni [1,2,3,4,5]. One popular type is the HO which is a half wave dipole folded into a square or a circle [6,7]. The impedance of this antenna is not 50 ohms and requires a matching section. If more gain is desired, then two or more elements can be stacked collinearly, with each element fed separately through power dividers.

A simpler way to combine elements is to connect the loops together by adding vertical extensions to the ends of each loop. The lengths of the loops and vertical wires are adjusted for maximum horizontal gain consistent with a 50 ohm feedpoint in the center of the bottom loop using an antenna modeling program. Since I decided to use 3/8 inch diameter soft copper tubing for the horizontal loops and # 18 wire for the vertical wires, I used mininec because nec cannot be used for antennas made of two wires of very different diameter [8,9]. It is possible to build this antenna with only one diameter wire, in which case nec can be used [10].

A test antenna was made for 144.5 MHz consisting of four horizontal loops made of 3/8 inch diameter soft copper tubing and two #18 (0.040 inch diameter)

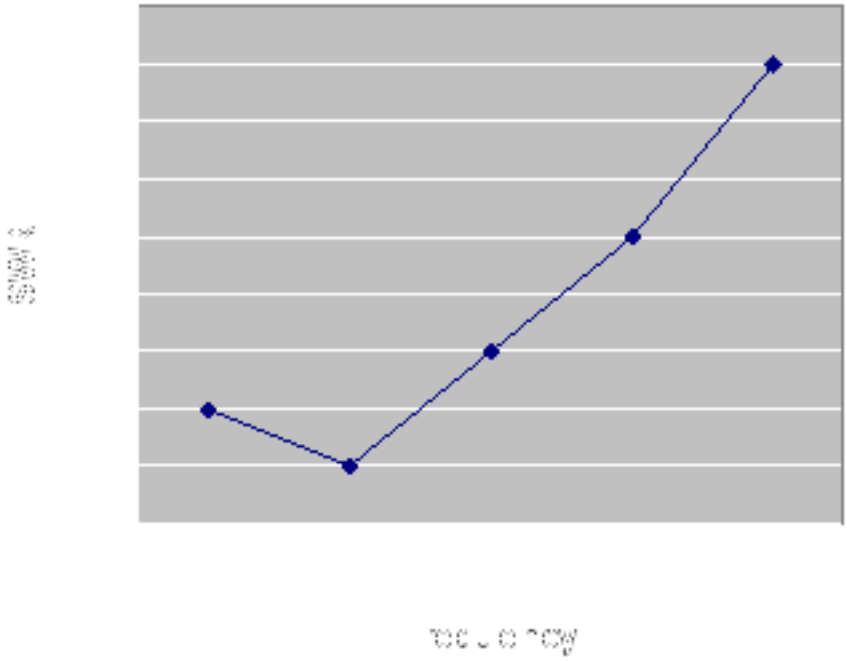
vertical wires. A loop drawing is shown in Figure 2. For the top and bottom loops, L = 31 7/16 and S = 1 5/8. For the middle loops, L = 28 5/16 and S = 1½. Cut the copper tubing to the lengths shown and bend them to approximate a circle with a gap as shown. All dimensions are in inches.



Then support the elements in some way. I used a piece of ¾ inch schedule 40 pvc pipe. The spacing between the elements (and the length of the vertical wires) should be 20 7/8 inches. Figure 3 shows how the vertical wires and the loops are connected together. This figure is drawn from the point of view off axis from the top of the antenna. The antenna is fed in the center of the bottom loop through a balun.

Figure 1 shows the assembled antenna on the roof. No patterns were taken, but the SWR (Figure 4) agreed with the simulated SWR (except for a 1% frequency offset error) which gives confidence that the antenna is working properly. The simulated gain is 3.27+/-1.22 dBi.

Figure 4



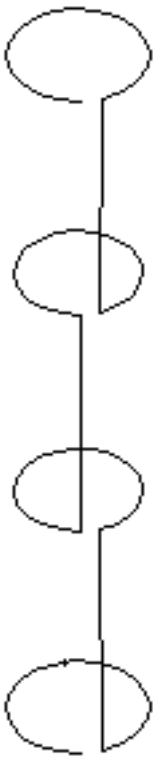


Figure 3

Adding more elements will improve the gain. If you decide to change the design using different diameters, more elements, or a different frequency, you should resimulate using mininec. Mininec

typically has frequency offset errors, so the final design may have to be tweaked a bit. I increased the vertical wire lengths from 20 to 20 7/8 inches from my original simulation (keeping the loop lengths the same). The frequency can also be adjusted by varying the gap at the end of the loops. Closing the gap lowers the resonant frequency (and also the impedance). Increasing the gap increases the gain variation (ie, makes it less omni). If you decide to use the same diameter for the vertical and horizontal elements, then nec can be used, but it may not be possible to get to 50 ohms.
One further comment. The spacing between the loops is smaller than is usually used in collinear arrays. This is because the electrical length between the centers of each loop is a half wavelength. The radiation from the loops is kept in phase by reversing the wire direction between adjacent loops (See Figure 3).

Ross Anderson W1HBQ May 21, 2004 ross_anderson@comcast.net

References and Links

[1] The Eggbeater EB-144
http://www.wb0w.com/m2/2mtr/2_meter_antennas.htm

[2] The Batwing http://www.qsl.net/kd2bd/batwing.html

[3] The Turnstile http://www.cebik.com/turns.html

[4] The Wheel http://www.hamtv.com/wheel.html

[5] Omniangles http://www.parelectronics.com/omnis.htm

[6] 2M HO Loop http://www.wb0w.com/m2/2mtr/2_meter_antennas.htm

[7] KB6KQ Loops http://www.kb6kq.com/products.html

[8] When MININEC is Superior to NEC
http://www.cebik.com/amod/amod56.html

[9] I use K6STI's AO. His e-mail address is bb at n2 dot net.

[10] I use 4nec2 by Arie http://www.qsl.net/wb6tpu/swindex.html

My homepage "Ross's Antennas" with links to my other pages is http://home.comcast.net/~ross_anderson

keyword: omnidirectional

Fan-tenna... An antenna made from a fan?

Did you ever look at a broken household item and think 'I can't throw this out, there just must be some use for it in my ham shack'! That was the case when our fan gave up the ghost one summer. The more I looked at the metal cage of that fan, the more it started to look like the perfect radial system for some future two meter ground plane antenna.

Some quick figuring (234/Frequency [MHz]), showed that if the cage radius is anywhere near 19 inches, it would make a 'fan-tastic' ground plane. As it turned out, my cage had a smaller diameter, in fact it was only about 10 inches, but I went ahead with the project anyway.

In an effort to create an easy to build, not to mention cheap, antenna, I decided to raid the hall closet to confiscate an old wire hanger to use as the vertical portion of the ground plane. I cut the wire a bit long to leave plenty of room for adjusting to a 1 to 1 SWR. With the help Bob (N3LSS), and his 2 meter hand held, we cut and trimmed, and ended up with a textbook 19.25" vertical radiator and a perfect match.

Construction:

Construction is extremely simple. First you'll need a connector, to properly mate with the one on the end of your coax. I used an SO-239 connector from my junk box. Insert the connector through the center of one side of the fan cage. My fan cage conveniently separated into two halves and provided easy access for mounting the connector.

If the center section of your cage is plastic, be sure to attach a jumper from the connector to the metal portion of the cage.
Solder the coat hanger to the connector.
Lastly, attach your coax after routing it through the back side of the other cage half.

If you like to tweak and prune, start with a vertical length of about 21 inches. Cut off only 1/4 - 1/2 inch at a time! If you don't have access to an SWR bridge for the two meter band, just cut the whip to 19 1/4 inches and your match will be very close.
If outdoor use is anticipated, I'd suggest using an aluminum rod in place of the coat hanger in order to survive the winds.
Although a 19 inch radius cage is optimum, as this design shows, a cage diameter even half that size works so don't bother to try and modify its length at all.

Pretty 'fan-cy' huh?

Test out:

But does it work? We are fortunate to have access to a local repeater which has a unique feature. It provides a voice announcement of your signal strength into the repeater. A quick test with the new fan-tenna indoors, in front of a window, and on the ground floor, provided a better than expected 30 over signal.
Additional tests from the same location, using the hand held's existing whip, proved that the new ground plane was far superior and provided an amazing improvement in gain. Yes indeed, this antenna does work!

Antenna Construction Tips
Back to the Antenna Elmer
Back to the Home Page

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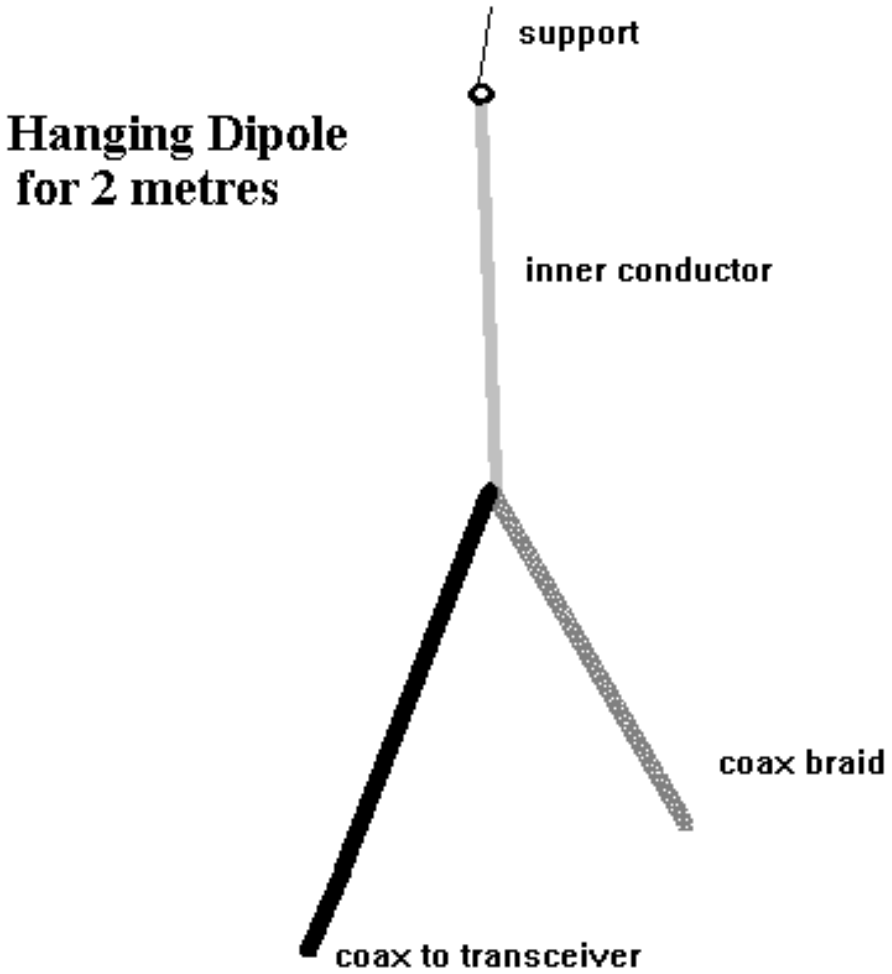
Build a hanging dipole for two metres

Described here is a simple omnidirectional, vertically-polarised dipole for two metres. Made from coaxial cable, it can be rolled up and stored in a small container. It may be used as is indoors, or waterproofed for use outside. No extravagant gain claims are made; this dipole has no more gain than any other. However, it should be significantly more effective than the antenna that came with your handheld. The cost of building the project is around five to ten dollars. Allow about 20 minutes to construct and erect the antenna.

A single length of 50 ohm coaxial cable forms both the antenna element and the feedline. The antenna is made by removing a quarter wavelength of outer jacket and bending the braid back along the cable towards the transceiver to form a vertical dipole. This means no metal work or wiring is required (apart from attaching the BNC or PL259 plug).

Download this image for a drawing of the completed antenna.



Parts required

The following is required to complete the project:-

- 3-4m RG58 coaxial cable (not critical - use longer length if height is needed or the operating position is distant from the antenna)
- PL259 or BNC plug (to suit transceiver)
- small metal lug, washer or nut
- tape measure, scissors, small screwdriver, long-nosed pliers, multimeter, fishing line, soldering iron

Construction

- Solder the PL259 or BNC plug to one end of the RG58 cable.
- From the other end of the cable remove 48 cm of the black plastic outer covering to expose the braid.
- With a small screwdriver (Phillips head is best) gently part the braid to make a small hole near where it ceases to be covered by the plastic jacket. Aim to make it about 5mm in diameter.
- Use either pliers or a screwdriver to pull the inner conductor out from inside the braid through the hole in the braid (Fig 2c).
- Fold the braid back along the cable towards the plug. Solder the end of the braid to prevent fraying.
- Remove about 5mm insulation from the inner conductor.
- Solder the end of the inner conductor to a small metal lug or nut.
- Thread fishing line through the lug or nut and hang the antenna in its desired position.

The antenna is now operational. You may wish to check the SWR and make it longer or shorter if the SWR is above about 1.5:1 at 147 MHz.

Erection and use

The antenna should be hung vertically for best performance. Keep it away from metal objects and have it as high as possible. Where signals are weak, hang the antenna near a window facing the repeater. If you intend to use the antenna outside, apply sealing compound to stop moisture entering the cable. Not doing this will mean poorer performance over time as cable losses increase.

---

back to Peter Parker's Project Page

This page was produced by Peter Parker VK3YE (ex VK1PK) (parkerp@alphalink.com.au). *The page is continually under construction. Material may be copied for personal or non-profit use only.*

The Grid Yagi

Figure 1

The Grid Yagi (or Grid Quad) is a high performance yagi antenna that can be built with readily obtainable inexpensive materials. Described here is a 6 element 2 meter version with a boom length of about 1 wavelength, shown in Figure 1.

The boom is made of 1½ inch pvc pipe, although any suitable material can be used, such as steel, aluminum, fiberglass, or wood. The elements are cut from 2 inch by 4 inch galvanized welded wire fencing, with a wire diameter of 0.078 inch, which is what #14 steel wire becomes when it is galvanized, and are shown in Figure 2. This fencing material and pvc pipe are available in any hardware store. The driven element and the four directors are all 24 inches by 24 inches. The reflector is 32 inches by 24 inches. The driven element has an 18 inch slot in it and is fed at the bottom of the slot. At the other end of the slot is a shorting wire.

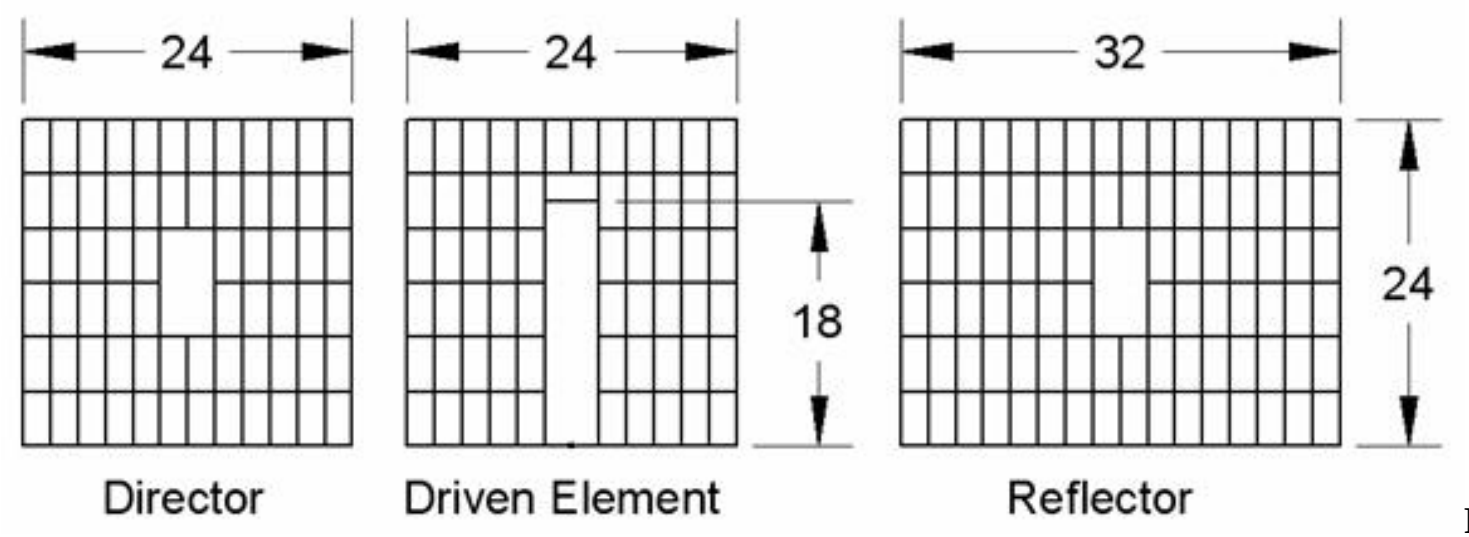


Figure 2

I attached the elements to the boom using 1/8 inch diameter fiberglass rods. Holes were drilled in the boom, and the rods passed through the holes and around the wires of the elements, two rods per element. Figure 3 shows the two fiberglass rods passing through the boom and around the wires of director D4. Rods were also used to stiffen the driven element where the feed slot was cut (not shown). The fiberglass rods were obtained locally at Tap Plastics [1]. It should be possible to use other stiff materials such as shish-kabob skewers. The element positions are shown in the Table. The elements can be glued in place or left unglued for easy disassembly.

Figure 3

Elements	Position
REF	0

DE	12 inches
D1	24
D2	38
D3	63
D4	86

Simulations were done using 4nec2, Arie's version of nec2, available free at Ray Anderson's Unofficial NEC Archives [2].

The antenna was fed with approximately 50 feet of RG-8 type coax with a measured loss of 1.3 dB through a W2DU type balun [3] with three Amidon beads, type FB-43-1020 [4]. Figure 4 shows the simulated and measured SWR. The SWR measurement was done using a MFJ-259B SWR Analyzer. I could have adjusted the length of the slot to move the SWR minimum to another place in the band if I desired.

The simulated gain is 11.9 dBi at 144 MHz, 12.1 dBi at 146 MHz and 12.3 dBi at 148 MHz.

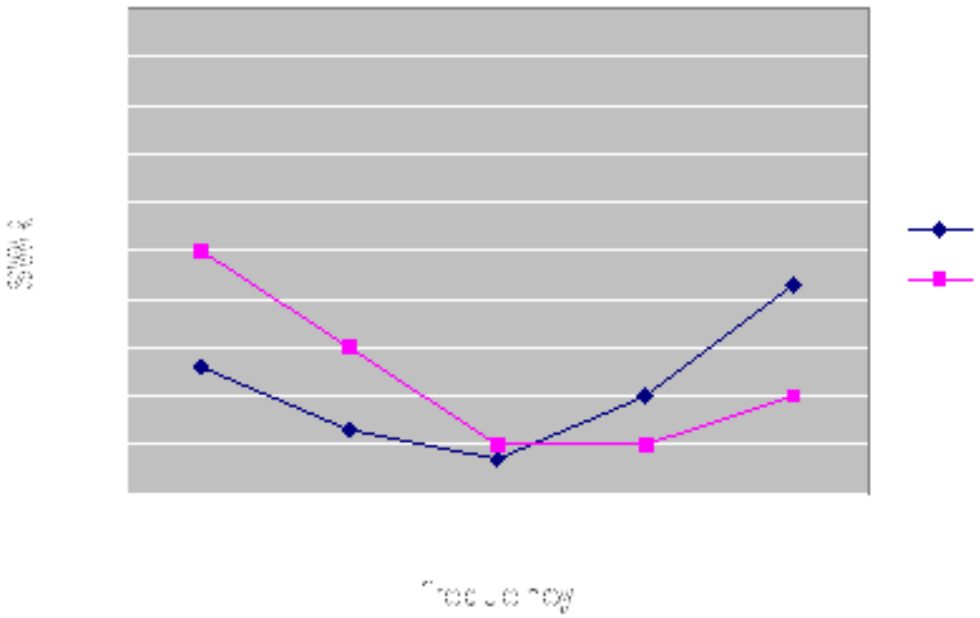


Figure 4

Ross Anderson W1HBQ February 15, 2004 ross_anderson@comcast.net

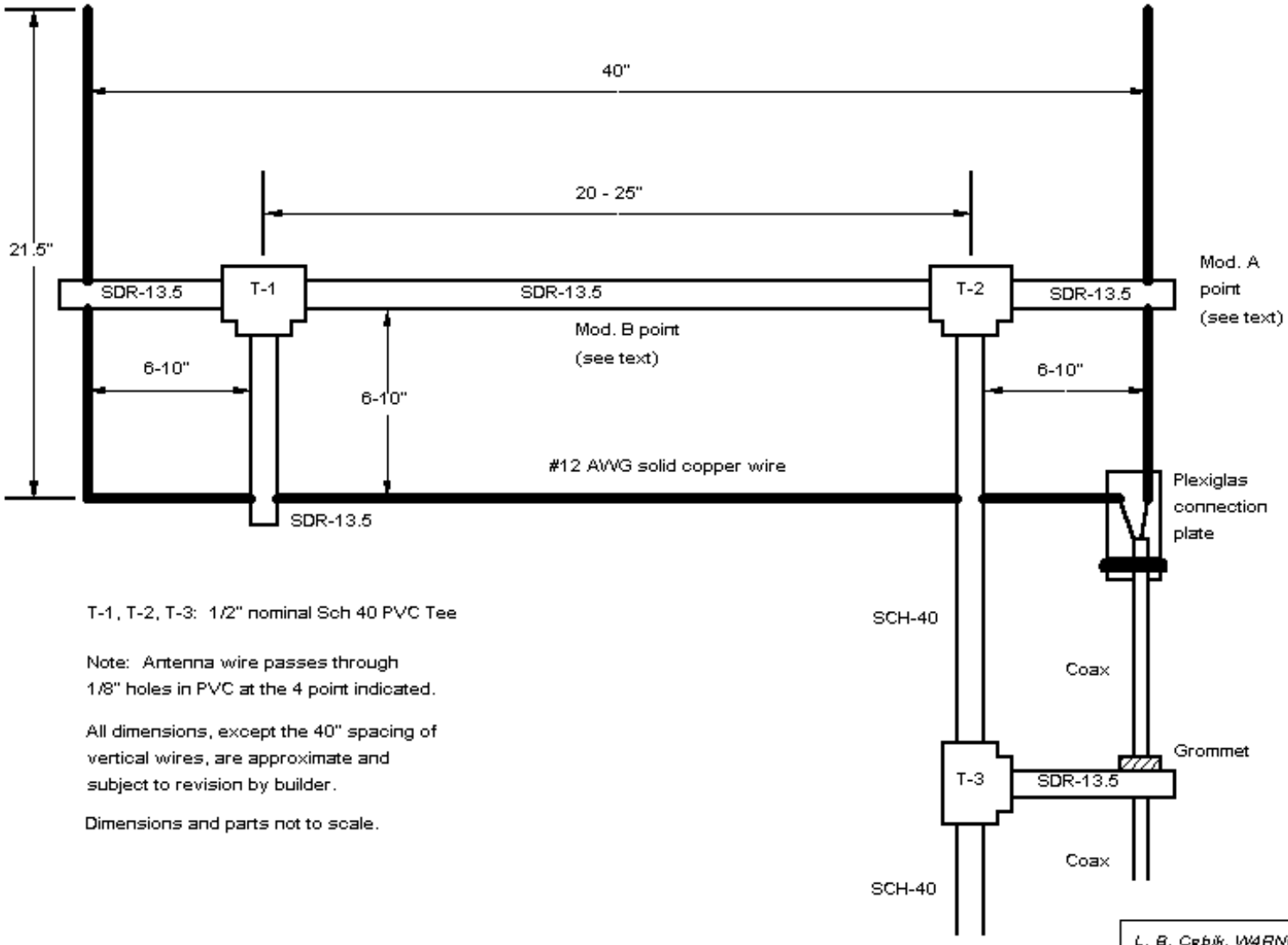
References and Links

[1] http://www.tapplastics.com/shop/product.php?pid=159
[2] http://www.qsl.net/wb6tpu/swindex.html

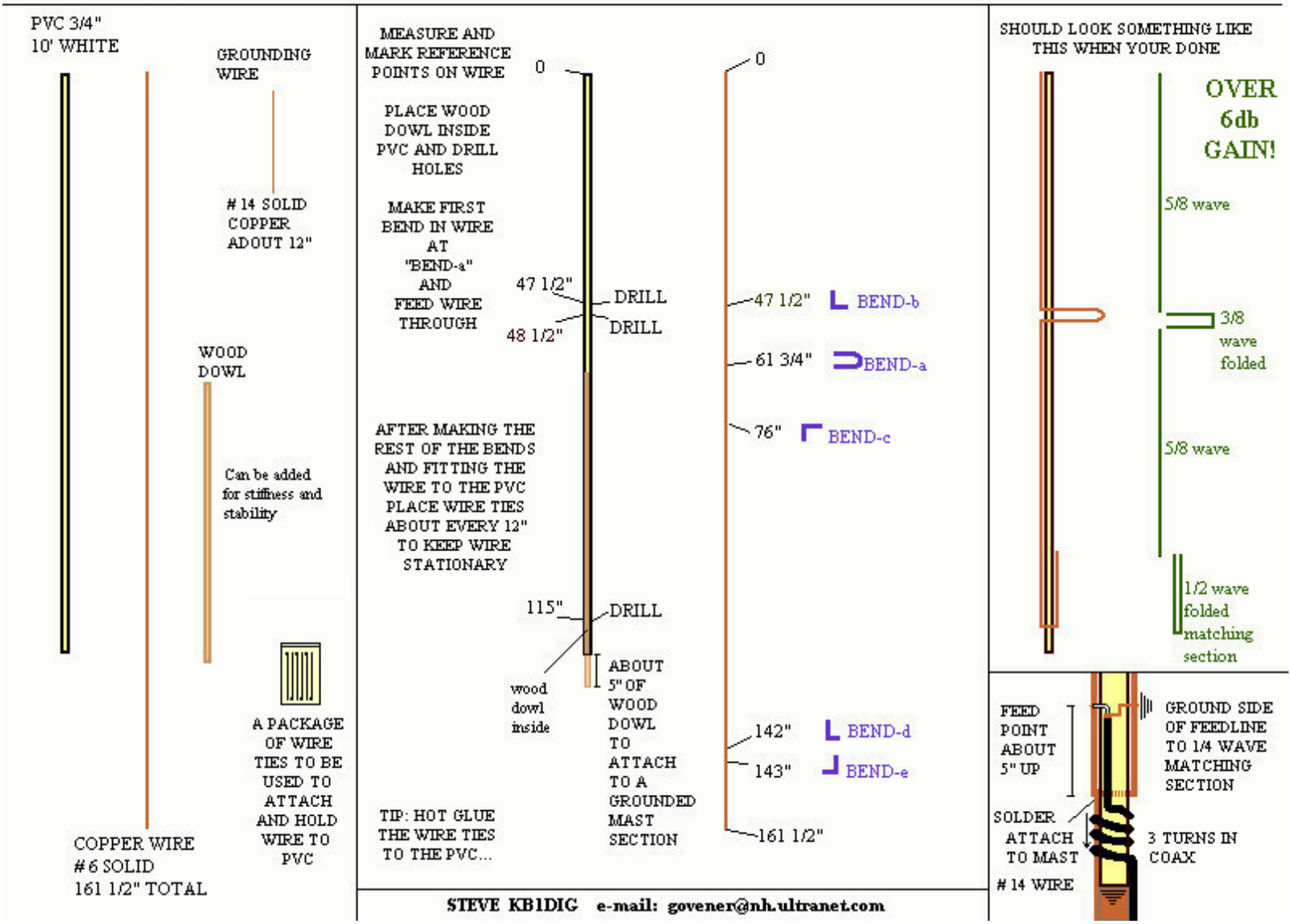
[3] http://home.iag.net/~w2du/Reflections2Chapter21SomeAspectsoftheBalunProblem.pdf
[4] http://www.amidoncorp.com/aai_ferritesshieldingbeads.htm

My homepage “Ross’s Antennas” with links to my other pages .
http://home.comcast.net/~ross_anderson

Keywords: disc yagi, loop yagi


40"
20 - 25"
21.5"
SDR-13.5
T-1
SDR-13.5
T-2
SDR-13.5
Mod. A
point
(see text)
Mod. B point
(see text)
6-10"
6-10"
6-10"
#12 AWG solid copper wire
SDR-13.5
Plexiglas
connection
plate
T-1, T-2, T-3: 1/2" nominal Sch 40 PVC Tee
SCH-40
Note: Antenna wire passes through
1/8" holes in PVC at the 4 point indicated.
Coax
All dimensions, except the 40" spacing of
vertical wires, are approximate and
subject to revision by builder.
Grommet
T-3
SDR-13.5
Dimensions and parts not to scale.
Coax
SCH-40
L. B. Cebik, W4RNL
Figure: 1
Title: Sketch of 2-Meter Half-Square PVC Support System
Project: 2-Meter Half Square
October 9, 1998

YOU CAN BUILD A 5/8th WAVE STACKED J-POLE FOR ABOUT $20.00!

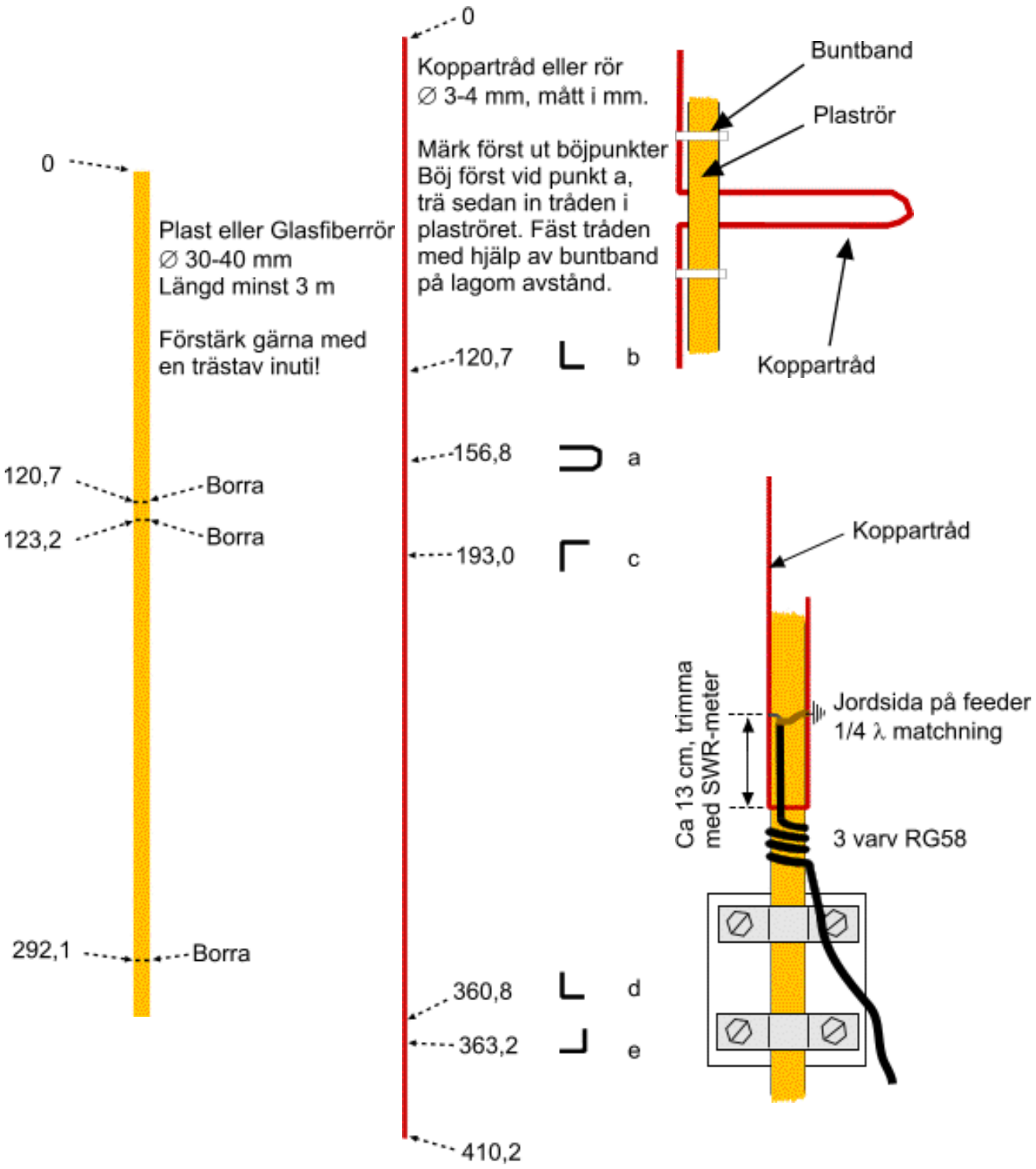


STEVE KB1DIG e-mail: govener@nh.ultranet.com

5/8 stackad Jpole för 2 m

Sidan finns på Spray
spray
Sponsrade länkar

Byggbeskrivning



Föregående sida ...

CHECK THIS OUT!

LED Projects!

Local ARES Web Project!

VHF ANTENNA Projects and Ideas:

2 METER ANTENNA DESIGN GALLERY

The KB1DIG & KB1GTR Ham Page

The J-pole is one of the most popular home-brew antenna designs for vertical 2-meter Amateur Radio. Another interesting home-brew project idea is the Halo design horizontal antenna. I did some experimenting with a few variations of these type of antennas and came up with the designs posted on this page. The main reason I considered building my own antennas was to gain a greater understanding of RF technologies and to save some money in the process. Each antenna has a unique use and are actually fun to build. Most of the parts can be found at your local hardware store. If you try one of these please drop me a line to let me know how it turned out.

Send E-mail to: buck0@comcast.net Good luck building! 73's Steve KB1DIG

Personal note: 11/21/01

Please understand that the intent of this new section is not to sound off or show off. This is the first addition I've made to this page in quite some time. I was worried about changing something because it has become so popular. Links to this antenna page have been turning up all over the Web. That's great! I've enjoyed answering all the e-mail. A lot of things have happened in the last few months that have inspired me to continue this new project. My wife, Kim, now has her Ham ticket! Her Call is KB1GTR and I'm very proud of her. We both had the opportunity to help with the NYC/WTC Relief Effort through the Salvation Army Team Emergency Response Network, or SATERN. We were working with Logistics Communications from Oct. 7th through Oct. 18th. Looking back on our experiences, we determined the need to expend our capability for emergencies on a daily basis. Aside from already being involved with local 2-Meters as Net Control Operators, we recently joined the New Hampshire ARES/ RACES.

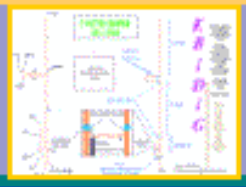
Thanks for all of your support! 73's Steve KB1DIG

we remember...the day the world changed

Personal note: 11/15/03 Please click here

Ladder Line J-Pole
2-Meter 1/2-Wave
1 Element

COPPER J-Pole
2-Meter 1/2-Wave
2 Element Collinear



INSIDE PVC J-Pole
2-Meter 1/2-Wave
2 Element Collinear

EMERGENCY J-Pole
2-Meter 1/2-Wave
2 Element Collinear

OUTSIDE PVC J-Pole
2-Meter 5/8-Wave
2 Element Collinear

"A Dedication to Emergency Service"

This is an off-site link to an article that appeared in the ARRL's QST Magazine, April 2002:
http://www.qsl.net/ka1ddb/adedicationtoemergencyservice.pdf

"A Dedication to Emergency Service" *Compiled and Edited by Rick Lindquist, N1RL.*

It tells our story, along with several other Amateurs that were there to help after 9-11-01.

Also, a link to the Mich-A-Con Amateur Radio Club main Webpage:
http://www.qsl.net/ka1ddb

EMERGENCY EQUIPMENT IDEAS

GO KIT: DAILY EMERGENCY COMMUNICATIONS GEAR IDEAS

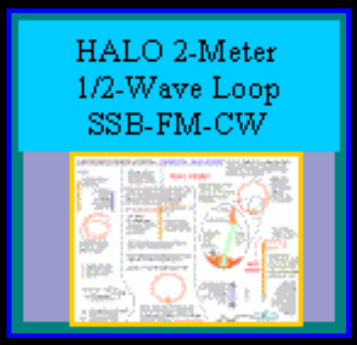
Nov. 21st, 2001, By Steve KB1DIG & Kim KB1GTR

How do we get all this HT equipment out of the drawer and into our hands when we need it? With a little research, we came up with the ideas and links posted in this section. Our favorite *quote* comes from C. Edward Harris, KE4SKY, VA RACES State Training Officer: *"It is better to have the bare essentials always handy than to leave a bulky pack someplace where you can't get to it."* This has been our inspiration for this project. As always, we are open to your suggestions and comments.

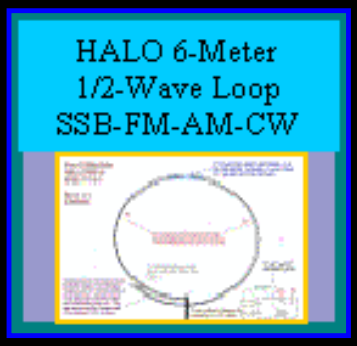
THE BOX: PORTABLE EMERGENCY COMMUNICATIONS STATION IDEAS

Jan. 9th, 2002, By Steve KB1DIG & Kim KB1GTR

Time for Plan B! This is a spin-off of the Go Kit project. Intended for portable long-term usage and also good for Field Day Events. Something requiring a little more power output than the HTs. A diamond in the rough. This "Box" is going to be under constant refinements this year with your suggestions and comments.



6 METER ANTENNA DESIGN



SOME ANTENNA WEB LINKS

- ARRL Web: Antenna Projects
- AC6V's HOMEBREW ANTENNAS LINKS
- Amateur Radio Information
- Antennas and ATV
- Jim's Notebook Index
- HAM UNIVERSE by Don N4UJW
- Amateur Radio Reference Library
- Harry's Homebrew Homepage
- W4ZT 2 Meter Dipole Antenna
- W4RNL
- Antenna Projects

Anderson Powerpole Ideas: Get Everyone Connected!

Nov. 6th, 2003, By Steve KB1DIG & Kim KB1GTR

Anderson Power Products and Powerpole® connectors. We have been doing research on this product and here are some of our findings. This overview may be familiar. Most of this information was found from different sources on the Internet and several Web links are listed on this page. We hope this will offer some new insight into how to go about integrating this Anderson Powerpole product into everyone's emergency power equipment needs. Our suggested attempts at leveling the playing field.. Also, keeping costs reasonable for the high-tech low budget practical minded Amateur in all of us. We are not experts and have no affiliation with the manufacturers or sales vendors mentioned. If you have something to contribute to this page and topic please send us your ideas and comments.

EMERGENCY LINKS

Amateur Radio Emergency Web Resources

RACES Click
Radio Amateur Civil Emergency Service

TAKE A BREAK !

To this end we present the intermission section.

- Technology for Country Folk
- You Know You're a Ham if
- NOTICE TO ALL VISITORS
- Little Dilbert and the Knack!

Our place to have fun and a place for you to enjoy.

Ham Radio Antennas

DXZONE.COM The DXZone DXZONE.COM

Rate this Site @ The DXZone

K1DWU
Ham Radio News, Links,
Classifieds and Much More

eHam.net
ham radio on the net

PLEASE CHECK OUT THESE HOME PAGES

Steve's and Kim's First Home Page!

Welcome to PatNJak.com

Equipment and Projects

ARES/RACES

Activation Plans
Projects
Local Nets
Events & Training
ARES Application
Net Control Script
EC Report

The following are some good emergency antenna projects to consider building. There are several reasons to consider building your own antennas for emergency work which include:

1. They are not expensive so if you loose them, or they get destroyed during the event you're not out big bucks!
2. You will gain a greater understanding of RF technologies especially antenna design.
3. You will learn how to improvise!
4. Last but not least, its fun!

Each antenna has a unique use. Please consider experimenting and playing! Share your success on the weekly nets as well!

Build a J-Pole for any frequency - Includes an Excel Spreadsheet!
Plans for a J-Pole

Pocket J-Pole
Half Wave J-Pole
Stacked 5/8
Field 2 meter Colinear
2 meter Colinear in PVC
2 meter Halo
Eggbeater Loop Antenna
HF Antenna Designs
Trickle Charge Circut

Off Site Plans

NVIS Short Range Antenna Systems for HF
NVIS Discussions
KB1DIG's Home Site
Copper Cactus Plans
N7QVC's J-Pole Plans
St. Louis Vertical (10-40m)
Windom - 6 bands in one antenna
Quickie Vertical- By KQ6RH
Wide Band Folded Dipole
Horizontally Oriented, Horizontally Polarized Large Wire Loop Antennas

Disaster Supply Kit

Being ready for an event or disaster requires preperation before the call comes. This means getting a supply kit together and knowing what is in it. We have included a article to help start your list.

Number of visitors:

n2njh@arrl.net
Buffalo, NY - USA
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This page was last updated on : undefined

Monday, September 06, 2004

The HAM Radio Operator's Antenna Handbook Page

another web site by;

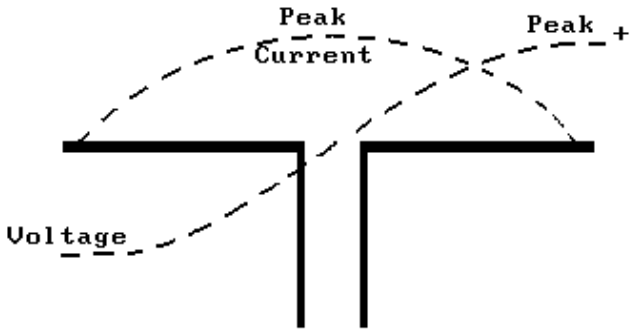
"Home of the RASCAL"
115 and 211 LUENBURG DRIVE; EVINGTON, VA 24550 FAX 434 525 7818

For TNC to Radio Interface Cable Diagrams, CLICK HERE !

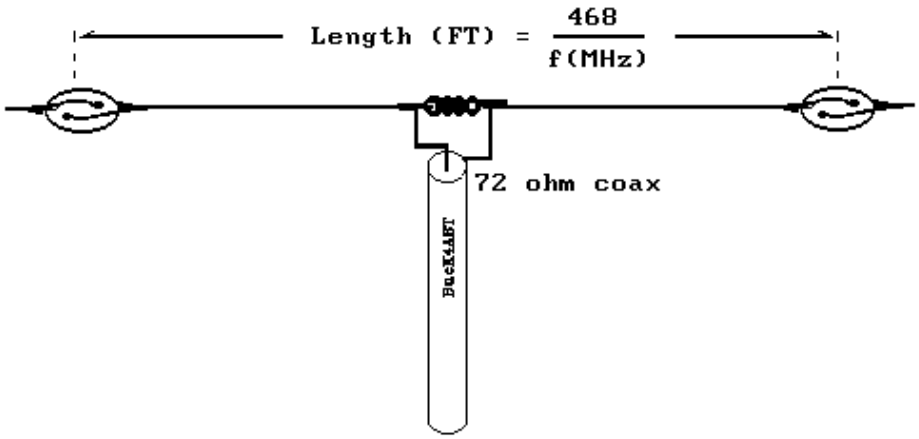
For PSK31 & SSTV Interface Diagrams, CLICK HERE !

- **Of all HF antennas, the multi-band WINDOM is my favorite.**
- **An improved 6 Meter "High-Performance" Base-Station Dipole.**
- **HF Antenna Cutting Chart**
- **VHF & UHF Antenna Cutting Chart**
- **Determine correct spacing between antennas to achive proper isolation.**
- **A two meter dipole**
- **It's very portable and EZ to build; A TV twin-lead two meter J-Pole**
- **Detailed drawing of a two meter "J" Pole antenna.**
- **Detailed drawing of a six meter "J" Pole antenna.**
- **Design your own "J" Pole antenna.**
- **Design your own QUAD antenna.**
- **A 4 element, six meter QUAD design using *Design-A-Quad*.**
- **Antennas and Antenna RF Meters**

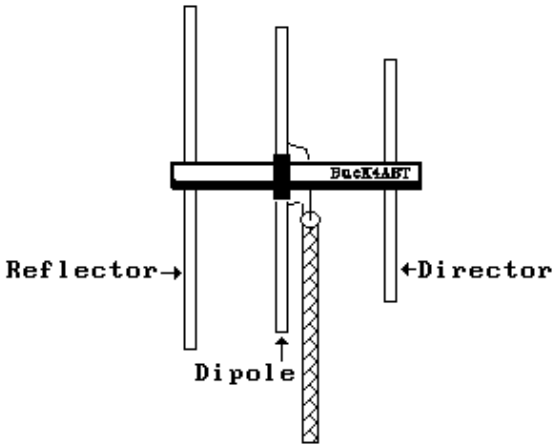
BUT FIRST.... let's review some antenna "basics" in pictorial form.



The basic dipole consists of two charges or elements which receive signals of opposite polarity.

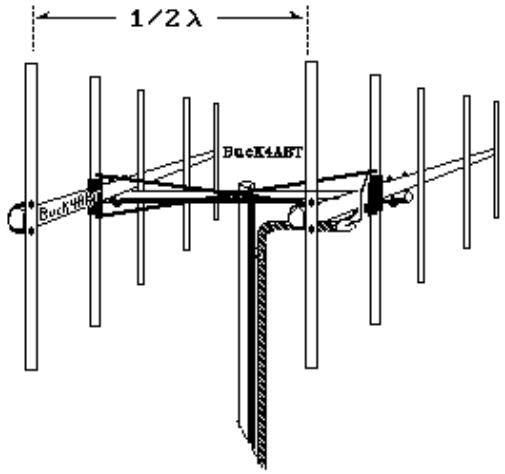


A dipole is a half-wave antenna fed at the center. The impedance at the center is near 72 ohms.

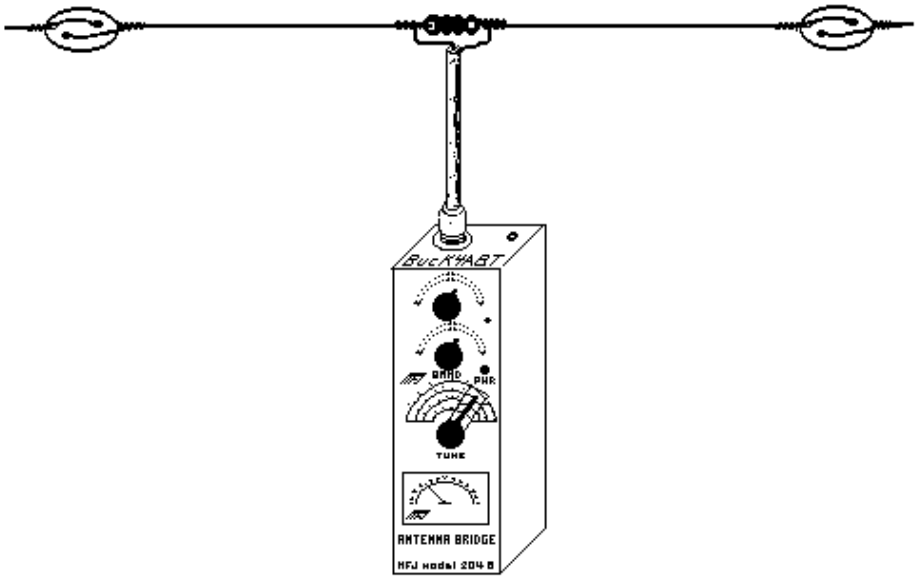


Adding a reflector and a parasitic element (director) to the dipole greatly improves its gain and directivity. Thus, the dipole becomes a "YAGI" antenna.

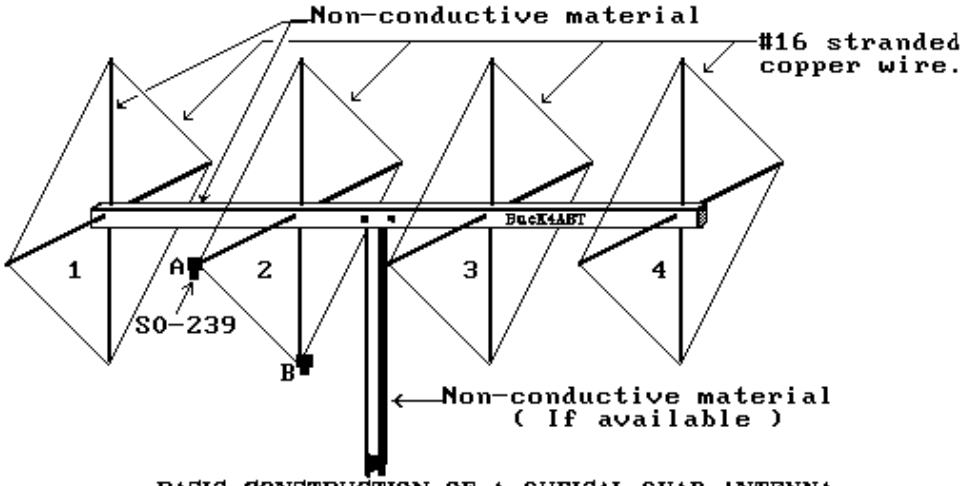
Good photos of VHF & UHF YAGI



Two yagis, stacked 1/2 wavelength apart will render another 3 dB of gain.

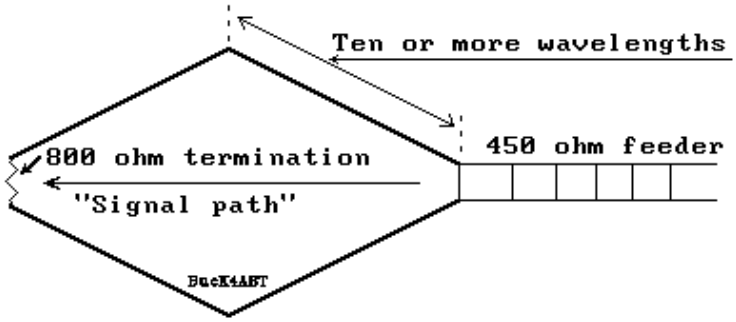


By using the "ANTENNA BRIDGE" the guess-work is removed and the correct cable length and impedance is sooner realized.

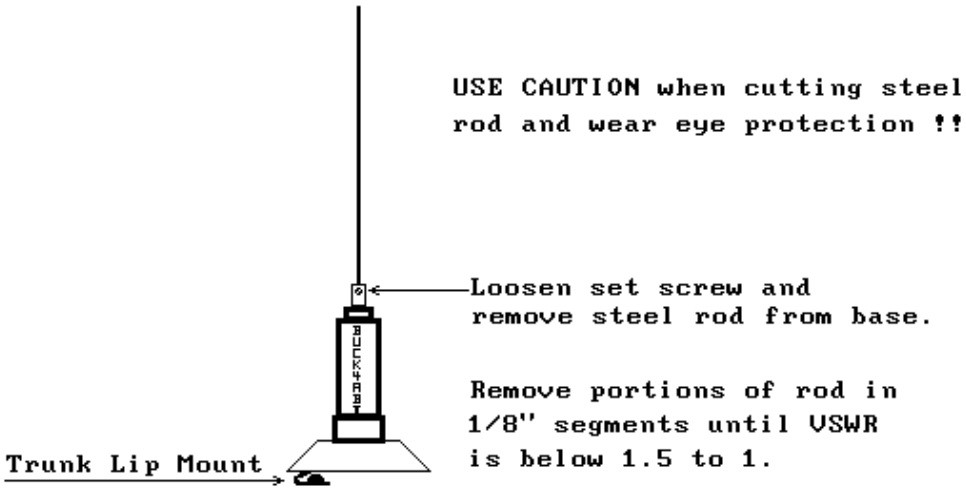


BASIC CONSTRUCTION OF A QUBICAL QUAD ANTENNA

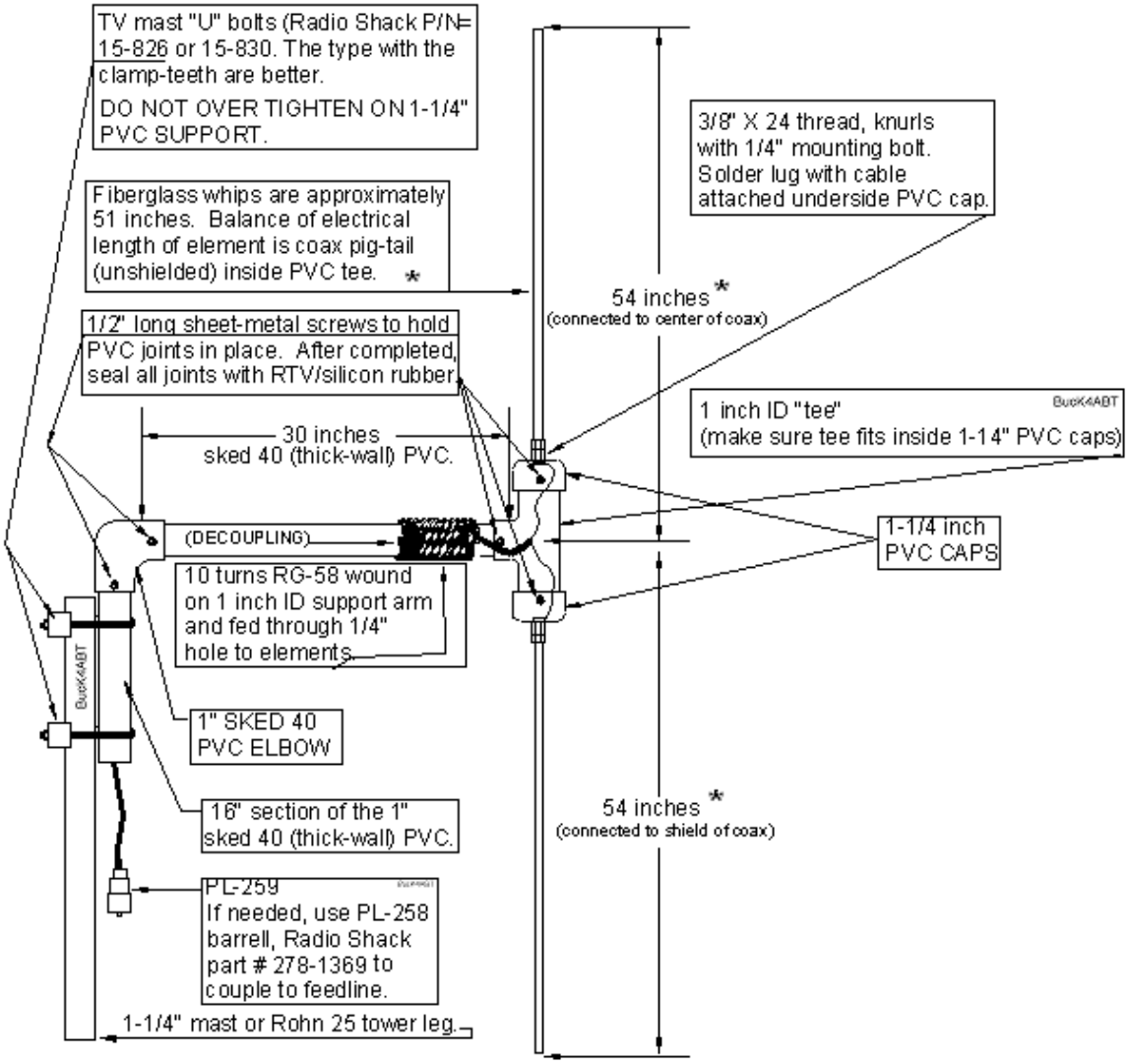
The "QUAD", feed point "A" for vertical polarization. Feed point "B" for horizontal polarization. Element (1) is reflector, (2) is driven element, 3 & 4 are directors.



The "ROMBIC" is a highly directional and very high gain antenna. It also requires a large parcel of real-estate.

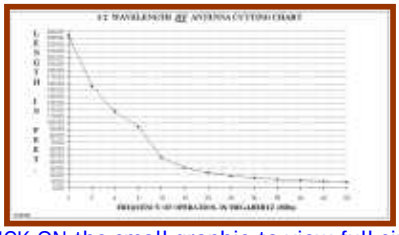


Method used to modify some CB antenna for ten meter use. NOTE: DO NOT over-tighten set screw.



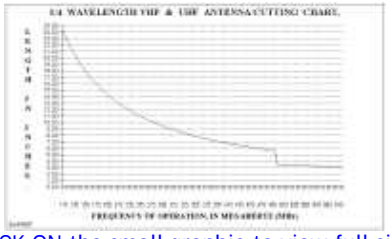
An improved "High-Performance" 6 meter Base-Station Dipole

Antennas & Accessories



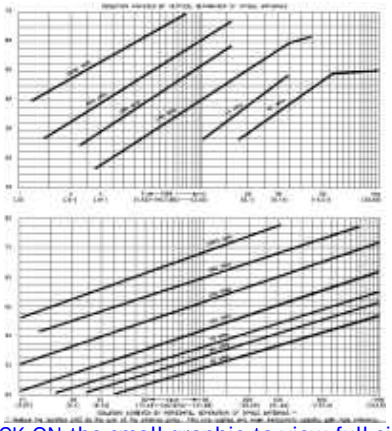
CLICK ON the small graphic to view full size.

Half-Wave HF Antenna Cutting Chart.



CLICK ON the small graphic to view full size.

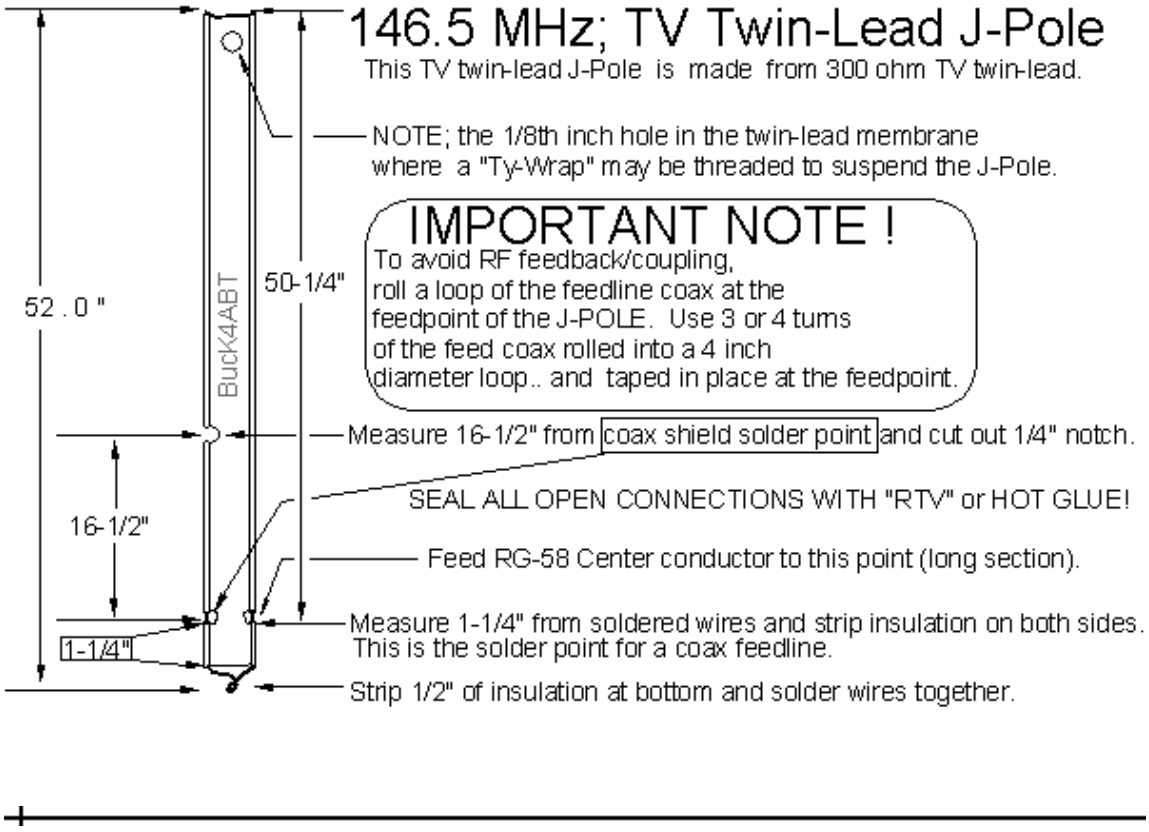
VHF/UHF Cutting Chart

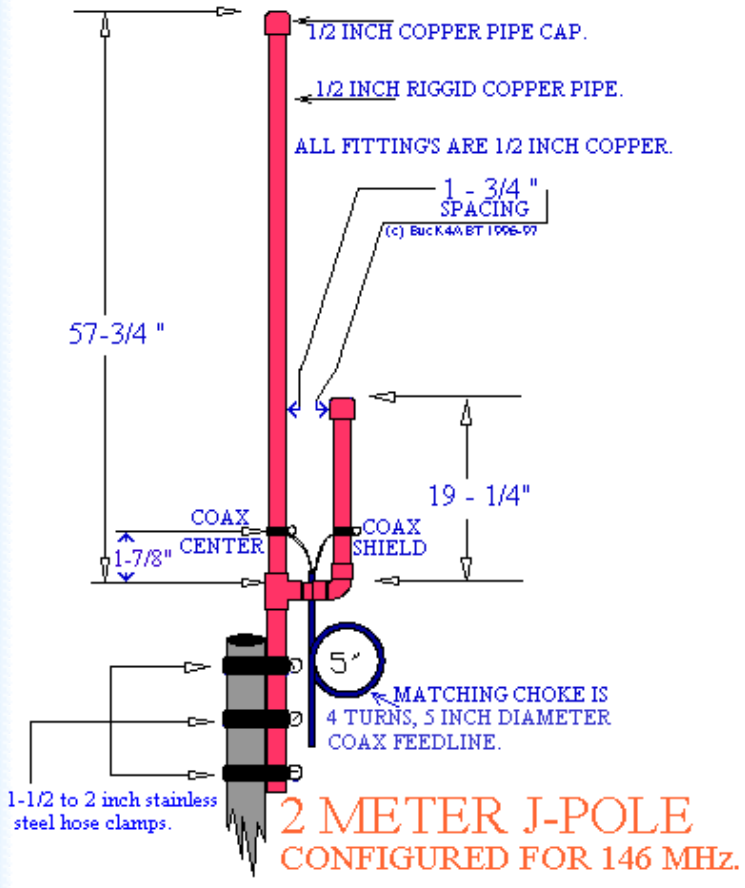


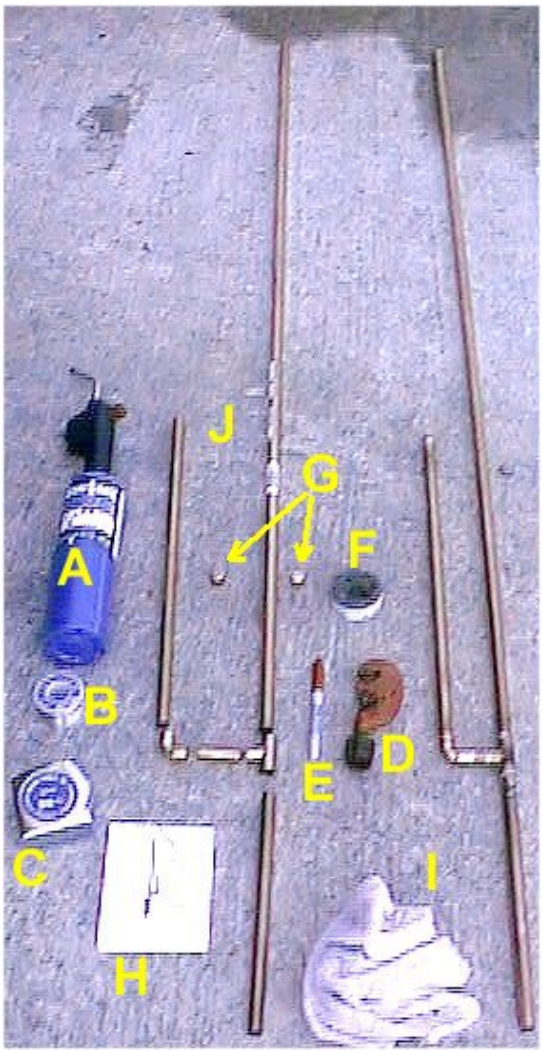
CLICK ON the small graphic to view full size.

Antenna Isolation / Spacing Charts

HAM Radio and Commercial 2-way radio antennas

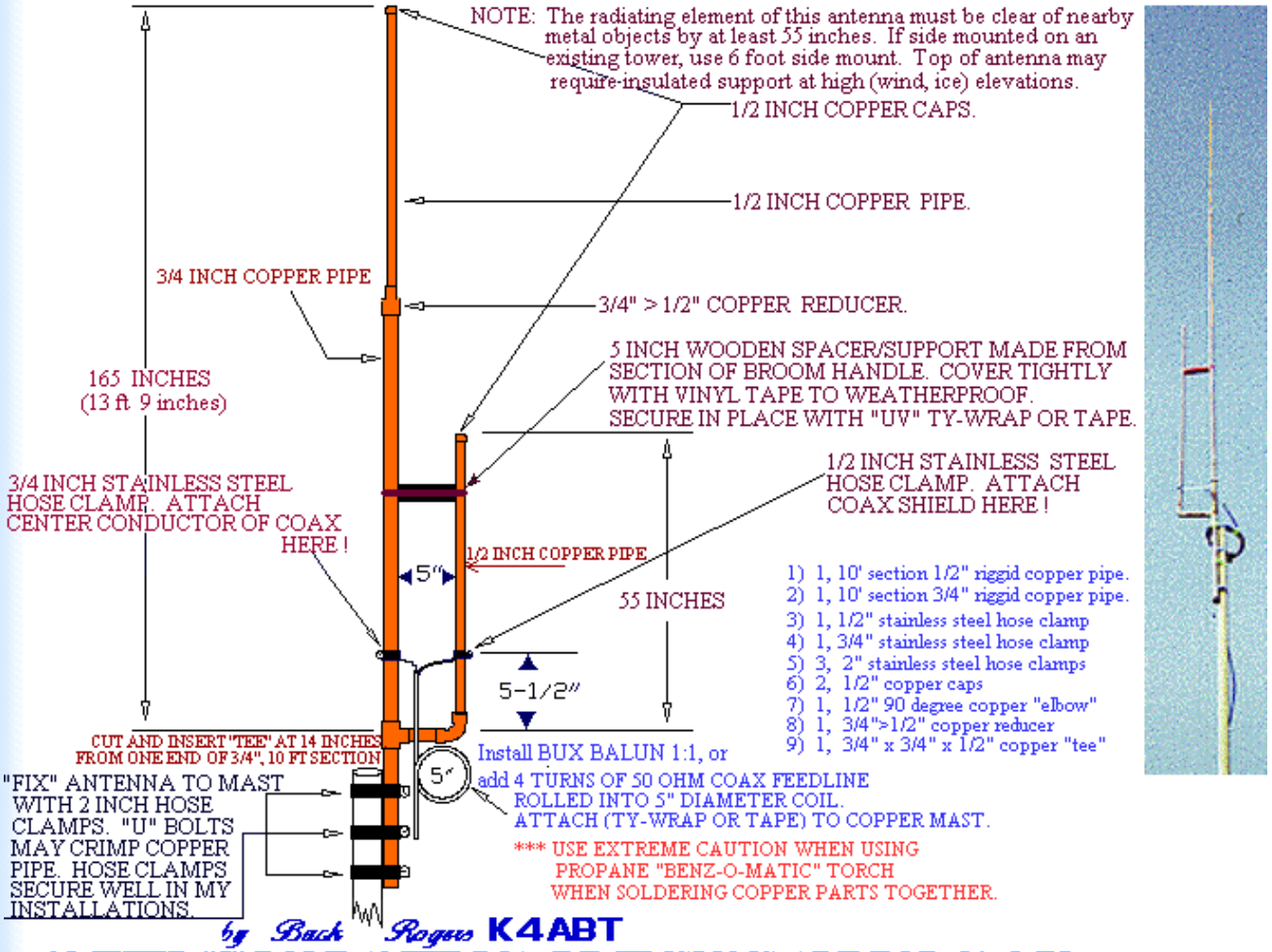


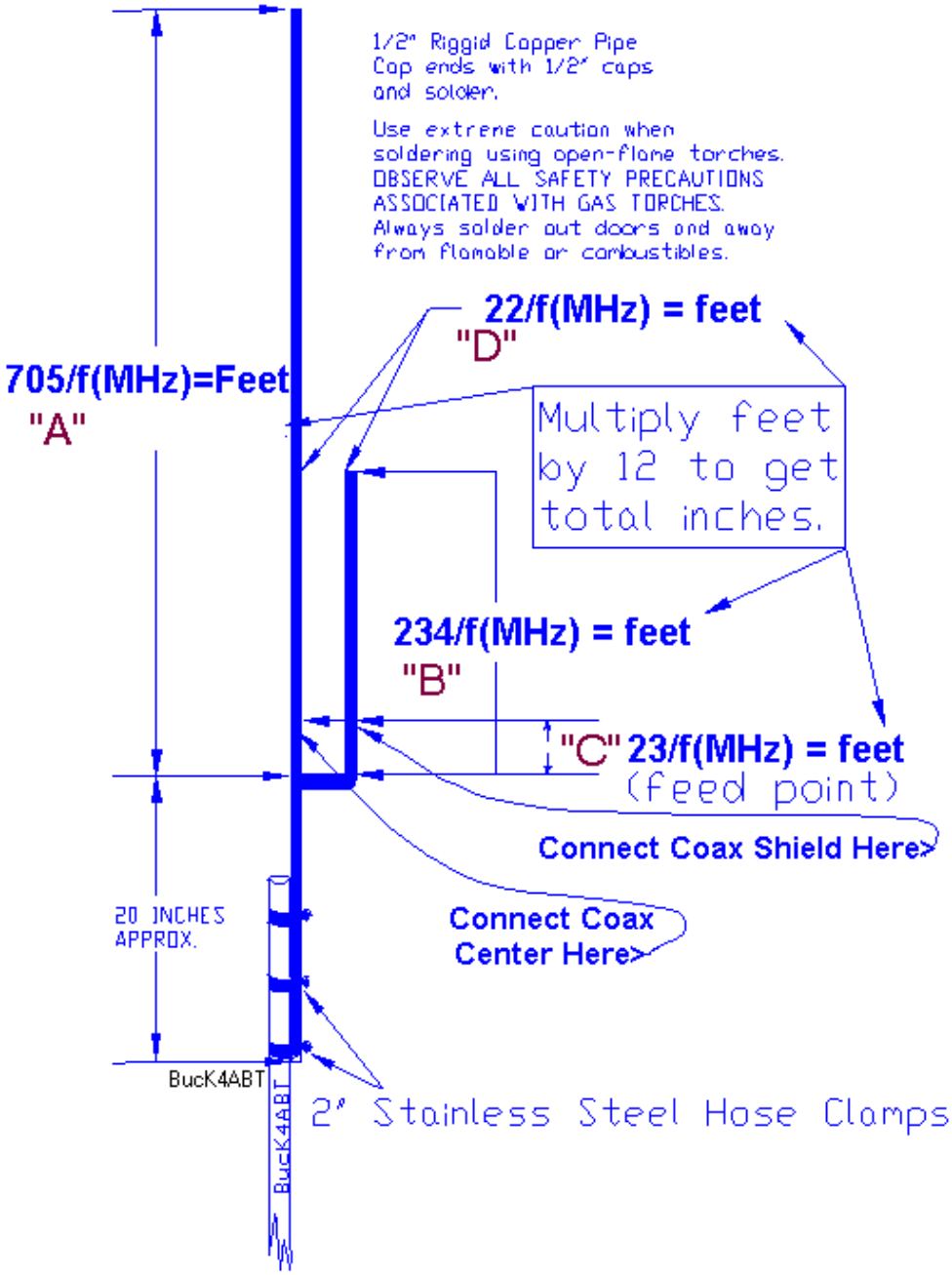




A=Benz-O-Matic propane torch; B=Lead-Free solder; C=Tape measure: D=Tubing cutter; E=Sharpie marking pen;

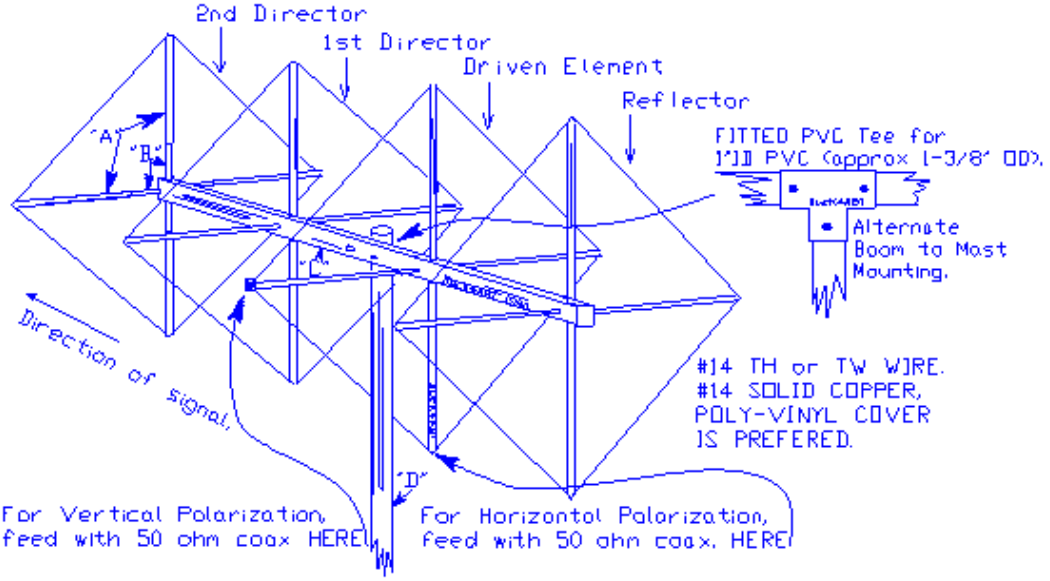
F=Solder Paste; G=1/2 inch copper caps; H=Hardcopy of the above drawing; I=Wet Towel; J=PreCut, ready to assemble parts of the 2 meter J-Pole.





CLICK HERE AND DESIGN YOUR OWN "J" Pole

Design-A-Quad Antenna



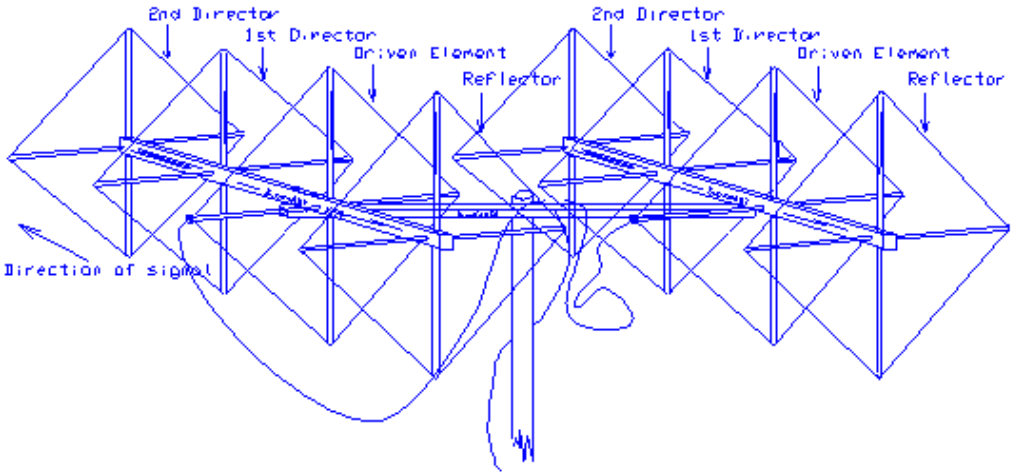
"A" = 1/2" ID hot/cold PVC. PVC should fit securely over 1/2" OD wooden dowel "B."

"B" = 1/2" OD wooden dowel 2 ft long through 1" ID boom. Dowel should extend approximately 9-1/2" out from each side of 1" ID boom. Fit 1/2" PVC ("A") over dowel. Make sure spreader is long enough to support each wire element length as determined by the calculations below. Once PVC is in place over the wooden dowels, drill with small bit and use short wire or picture hanger nail to secure on the dowel. Hot glue may be used as an additional keeper over the wire or nail head.

"C" = Boom may be made from square fiberglass or from round PVC 1" ID. Boom is drilled with 1/2" bit at points where wooden dowels are to be inserted.

NOTE: Make sure the vertical and horizontal dowel holes are drilled offset so the dowels will by-pass each other inside the boom. Drill dowels with 1/16" (or smaller) bit each side of boom and secure with small nail or wire to hold (centered) in place. BOOM is thick wall "Sked 40" 1" ID PVC.

"D" = 1" ID (sked 40/Thick wall) PVC 48" long. Use "fitted" wooden dowel inside to make it more rigid and provide "body" to area where "U" bolts may be attached (rotor head and to boom). See alternate boom mounting.



Horizontal "stacking" of quads will improve the signal strength by at least 3 DB. The QUAD's inherent immunity to terrestrial noise, makes it the ideal antenna (beam) for Packet Radio use.

Measurements and dimensions for the QUAD antenna are found using the QUAD design calculator below.

Calculate the total wire length for each element of a quad antenna.

REFLECTOR = 1030 / F (MHz)

DRIVEN = 1005 / F (MHz)

DIRECTOR 1 = 975 / F (MHz)

DIRECTOR 2 = 960 / F (MHz)

ELEMENT SPACING ***between:***

1) Reflector & Driven Element = 730 / F (MHz)
2) Driven Element & 1st Director= 600 / F (MHz)
3) 1st Director & 2nd Director = 600 / F (MHz)

Enter the formula for the antenna calculation

Divided by Freq MHz

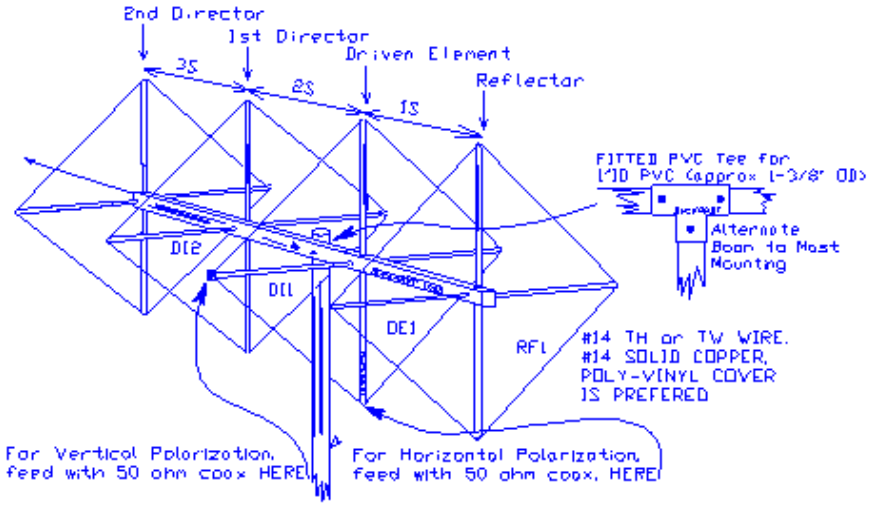
Total wire-length in feet is ***FEET.***

Total wire-length in inches is ***INCHES.***

One of four sides in FEET = **<- Use ONLY this number for "element spacing" when entering the *element spacing* numbers (constants; 730 or 600) above!**

Total all spacing measurements (in inches) then add 2 inches for each end.

THE QUAD SHOWN BELOW IS AN EXAMPLE OF A QUAD DESIGN USING *DESIGN-A-QUAD* ABOVE



AN EXAMPLE OF A 4 ELEMENT, SIX METER QUAD

TOTAL ELEMENT WIRE LENGTHS	LENGTH OF ONE SIDE
RF1 = 241-3/4 INCHES	60-1/2 INCHES
DE1 = 236-0"	59-.0"
DI1 = 229-.0"	57-1/4"
DI2 = 225-.0"	56-1/4"

ELEMENT SPACING
1S RF1 to DE1 = 43 INCHES
2S DE1 to DI1 = 34 INCHES
3S DI1 to DI2 = 34 INCHES
ADD 2 INCHES AT EACH END
TOTAL BOOM LENGTH = 115 INCHES

When stacking two of these quads, use 44 inch lengths of 75 OHM cable from Coax "TEE" connector to each Quad. Coax from transmitter to "Tee" is 50 ohm coax. As viewed above, for vertical polarization, feed both quads as shown (left side). For Horizontal polarization, feed both quads at bottom of the driven element (as indicated).

ANTENNA INSULATORS

Weatherproof, (TENYTE) insulators. Perfect for your DIPOLE or Windom antenna.

For the apartment dweller, you can now hang the 20 meter doublet in the attic. I've QSO'd with stations all over the world with the 33 ft dipole in the attic of my garage. One insulator at the center, and one at each end.

It's great for other HF WINDOM, single, and multi-band dipole antennas. Dielectric strength is superior to the old ceramic insulators, without susceptibility to cracking or breakage under impact or extreme temperature changes.

Package of 3, $2.99 ANTINSL3 Order Now

Package of 10, "TeNyte" BUX TNT Insulators $8.99, ANTINSLX10 Order Now

For those who want lightning protection, here is one of many, lightning protection device.

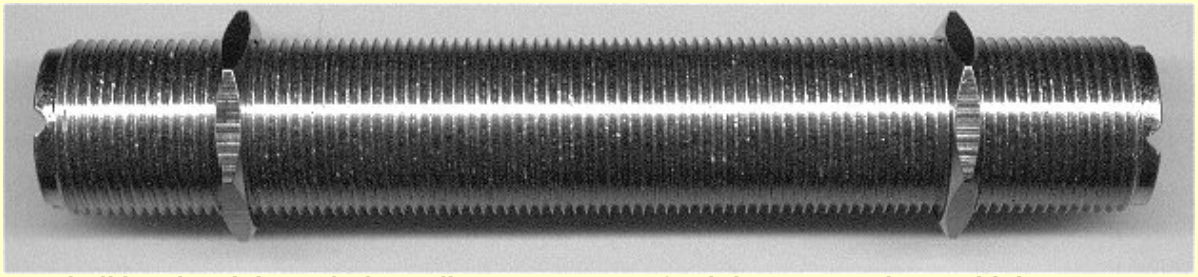
A Secondary Lightning Protector

Remember the axiom; ***"An ounce of prevention is worth a pound of cure."*** This lightning protector contains a modified spark gap that provides that ounce of prevention. We offer this device as a *secondary* radio protector. Remember, there is no substitute for a good earth ground, AC line surge, and spike protection.

Keep all ground leads as short as possible. Where lightning is involved, we make no warranties.

$ 3.95 each 7516 Order Now

UHF DOUBLE FEMALE BULKHEAD (feed-thru) CONNECTORS



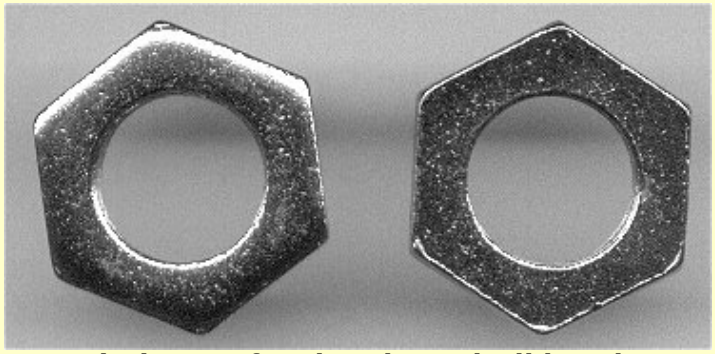
For bulkhead and through-the-wall UHF connector feed-thru connections, with keeper nuts.:

Order 7518-2 (Two inches)	**$ 2.95 ea.**	Order Now	**10 for $ 24.50**	Order Now
Order 7518-4 (Four inches)	**$ 3.95 ea.**	Order Now	**10 for $ 34.50**	Order Now
Order 7518-6 (Six inches)	**$ 4.95 ea.**	Order Now	**10 for $ 44.50**	Order Now

Order 7518-8 (Eight inches) **$ 8.95 ea.** Order Now **10 for $ 82.50** Order Now

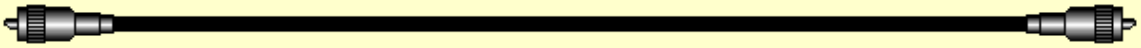
Order 7518-10 (Ten inches) **$ 10.95ea.** Order Now **10 for $ 99.50** Order Now

Order 7518-12 (Twelve inches) **$12.95ea.** Order Now **10 for $ 114.50** Order Now



Heavy Duty (1") Nuts for the above bulkhead connectors.

2/.99 cents HDM1-NUT Order Now **10 for $ 3.99** Order Now



50 Ohm impedance, 3 ft, 6 ft, 9 ft, 12 ft, & 18 feet.

These jumpers are made from high quality, low-loss, RG8X coax cable WxTite, PL-259 connectors installed at each end.

Three ft. (3') **ORDER No. 8X3-PLPL $3.95** Order Now

Six ft. (6') **ORDER No. 8X6-PLPL $4.95** Order Now

Nine ft. (9') **ORDER No. 8X9-PLPL $5.95** Order Now

Twelve ft. (12') **ORDER No. 8X12-PLPL $7.95** Order Now

Eighteen ft. (18') **ORDER No. 8X18-PLPL $8.95** Order Now

For more RF Connectors and communications site accessories; CLICK HERE !

At BUX COMM, *We don't cut corners!

The components used in the manufacture of our BALUNs are from top quality components, beginning with the Silver Plate SO239 connectors. The center insulator is teflon™(E.I Dupont). The wire we use to wind the ferrite donut is heavy-duty, silver flashed wire, with teflon™ insulation that will handle RF voltages above 5000 volts, and temperatures above 2000 degrees. The binding posts are heavy-duty, tempered brass, with end holes and side-thru holes to accommodate either type loop-thru connection. A double-shoulder brass capture nut is used to add a secure bite and improve antenna wire electrical connections.

4:1 Balun, BUXBALUN 41 $19.95 Order Now

o 50 ohm, SO-239 unbalanced input
o Balanced output
o 1.6 to 50 MHz
o Toroid (Voltage) design
o Heavy Duty, Lightweight construction
o Sealed against moisture

1:1 Balun, BUXBALUN 11 $19.95 Order Now

o 50 ohm, SO-239 unbalanced input
o Balanced output
o 1.6 to 50 MHz
o Toroid (Current) design
o Heavy Duty, Lightweight construction
o Sealed against moisture

BUX UN UN *De-Coupling transformer, similar to above,* but has *SO-239 (female) input* connector and *output connector is 1 ft Mini 8 cable with PL-259 (male).*

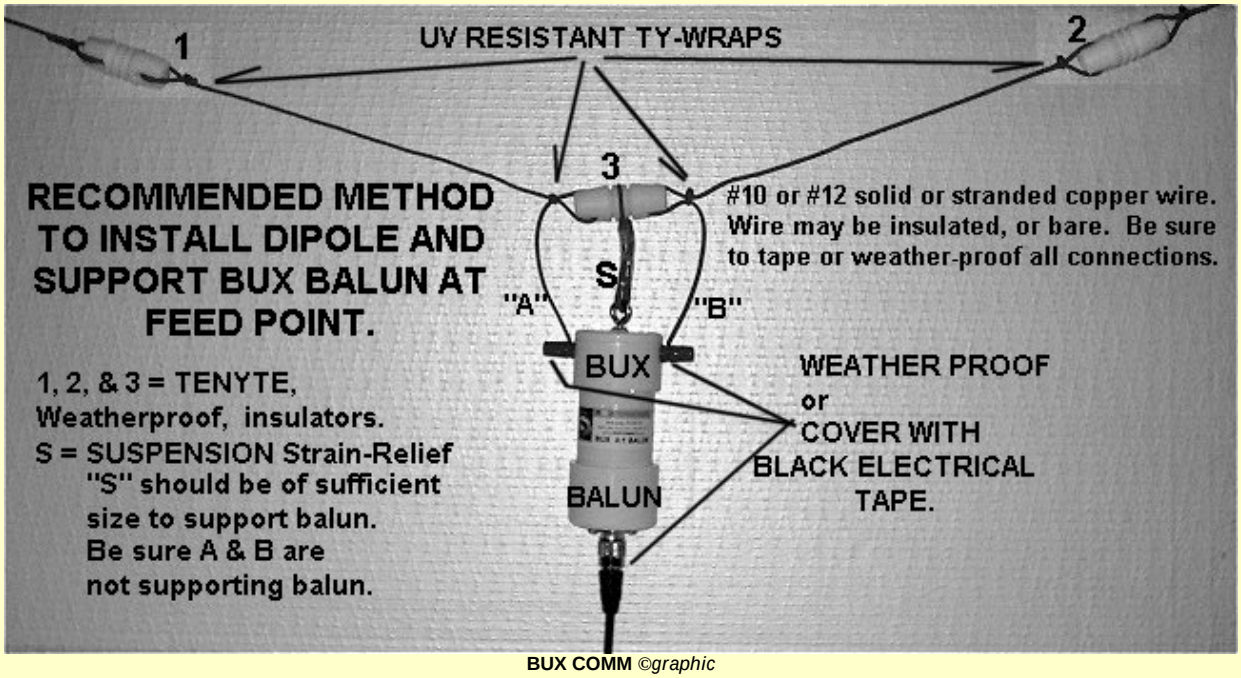
BUX ONION $19.95 Order Now

4:1, *1.5kw* Balun, BUX BALUN 41HD $27.95 Order Now

Toroid design, wound with teflon covered, silver plated wire.* Heavy-Duty, construction.

1:1, *1.5kw* Balun, BUX BALUN 11HD $27.95 Order Now

Toroid design, wound with teflon covered, silver plated wire.* Heavy-Duty, construction.



BUX COMM ©*graphic*

BUX MOBALUN

When you hear those strong HF mobile signals, here's one reason they stand out from the rest.

By installing the BUX MoBalun near the input to your antenna, you deliver more RF energy to the antenna. At the same time, the BUX MoBalun prevents RF from traveling back along the shield (high SWR) of your coax. High power rating, Low-permeability toroid, with Internal composition fiber-glass to prevent vibration during mobile operation. For input and output connectors, we use only the best Silver plated, Teflon insulated, SO239 connectors. 500 watts PEP. ***This MoBALUN is also ideal for marine antenna installations.***

BUX COMM ©*Photo*

$19.95 MOBALUN Order Now

CLICK HERE for Accessories, Antennas, Components, Connectors & Power cords

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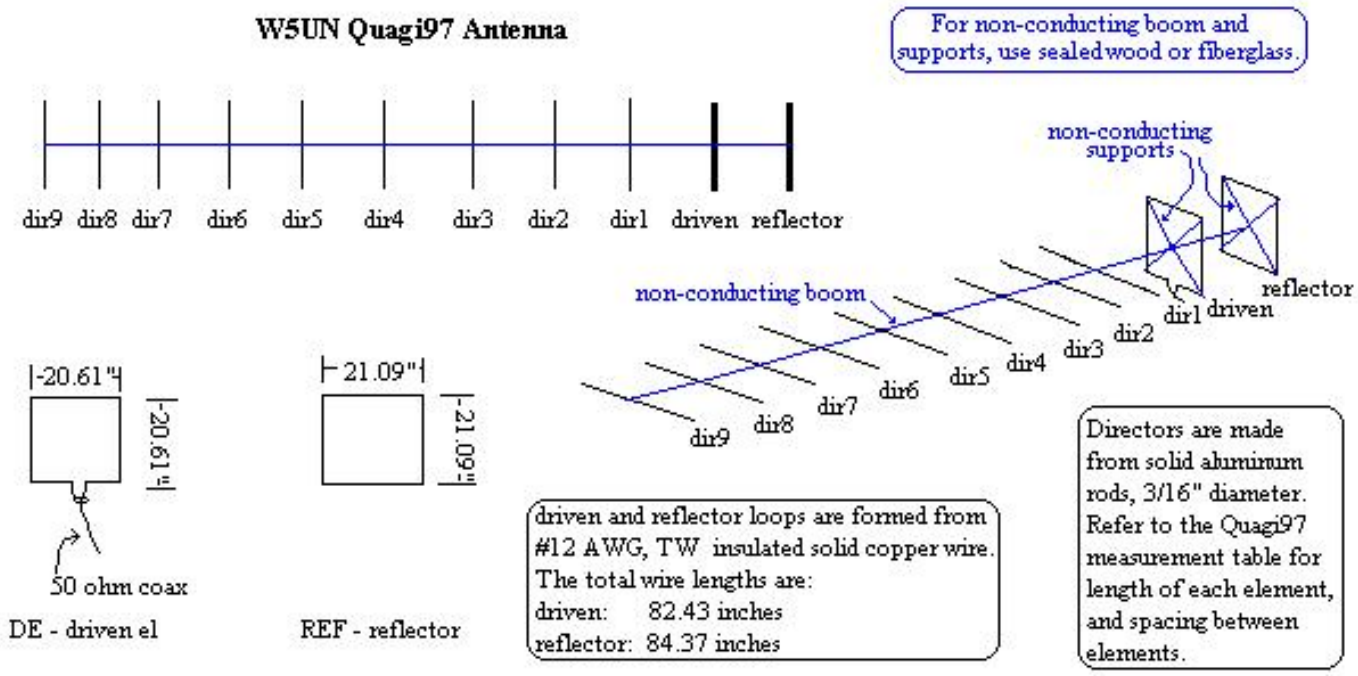
FAX 434 525 7818

THE W5UN 2 METER QUAGI 97
Issue 2, July 17, 1997

The W5UN Quagi 97 is a computer optimized quagi derived from the original W5UN Quagi of the early 1980s. The length of the 1997 model has been deliberately shortened to allow it fit on a 24 foot boom. Performance is substantially higher than that of the original antenna. It now compares favorably with yagis of the same length. The two programs used to optimize the updated antenna were AO6 and NEC 2.

This antenna has not been verified by actual construction. Gain and front to back will be as stated below. The only component of the antenna that needs verification is the 50 ohm match. *Who would like to be the FIRST to build the actual prototype model??* Full consultation will be available to the prototype builder during construction. Contact w5un@wt.net if you wish to be the first!



Antenna Characteristics:
23' 9"
11 Element Gain: 13.56 dBd at 144.100 Mhz
F/B : 23 dB Stacking: Optimum: E Plane: 13.73'
Optimum: H Plane: 12.86'
Stacking can be reduced up to 90 percent of optimum and still achieve acceptable stacking gain. Optimum stacking is recommended, however. W5UN QUAGI 97 Dimensions in FREE SPACE:

Element	Boom Pos.	El. Length	Material
Reflector	0.0000"	84.37" loop.	#12 solid INSULATED copper wire *
Driven	16.2500"	82.43" loop.	#12 solid INSULATED copper wire *
Dir 1	37.1875"	36.168"	3/16" aluminum rod
Dir 2	70.1875"	36.052"	3/16" aluminum rod
Dir 3	102.5000"	35.817"	3/16" aluminum rod
Dir 4	135.8125"	35.475"	3/16" aluminum rod
Dir 5	169.0625"	35.179"	3/16" aluminum rod
Dir 6	202.8125"	35.090"	3/16" aluminum rod
Dir 7	234.2500"	35.324"	3/16" aluminum rod
Dir 8	264.8750"	35.617"	3/16" aluminum rod

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Dir 9      285.0000"     35.108"       3/16" aluminum rod
```

* #12 TW insulated solid copper wire is common house wire, found at most hardware and building supply stores. Do not strip the insulation.

The above dimensions may be used for non metallic booms like wood or fiberglass.

If you wish to use a metal boom with through the boom insulated elements, please apply the following correction to the directors only. The reflector and driven quad elements must remain insulated from the boom.

Boom Diameter	Correction	Add
0.750" or 19.050MM	10.56%	.0792" or 2.02MM
0.875" or 22.225MM	12.14%	.1062" or 2.70MM
1.000" or 25.400MM	13.66%	.1366" or 3.47MM
1.125 or 28.575MM	15.13%	.1702" or 4.32MM
1.250" or 31.750MM	16.54%	.2068" or 5.25MM
1.375" or 34.925MM	17.90%	.2462" or 6.25MM
1.500" or 38.100MM	19.21%	.2882" or 7.32MM
1.750" or 44.450MM	21.67%	.3792" or 9.63MM
2.000" or 50.800MM	23.91%	.4783" or 12.15MM

(End of Table)

Further information concerning use of metallic booms with insulated 'thru-the boom' mounted elements follows:

PROVEN ACCURATE FOR BOOM DIAMETERS SMALLER THAN .055 WAVELENGTHS.
MEASUREMENTS BY DL6WU.FORMULA BY G3SEK.

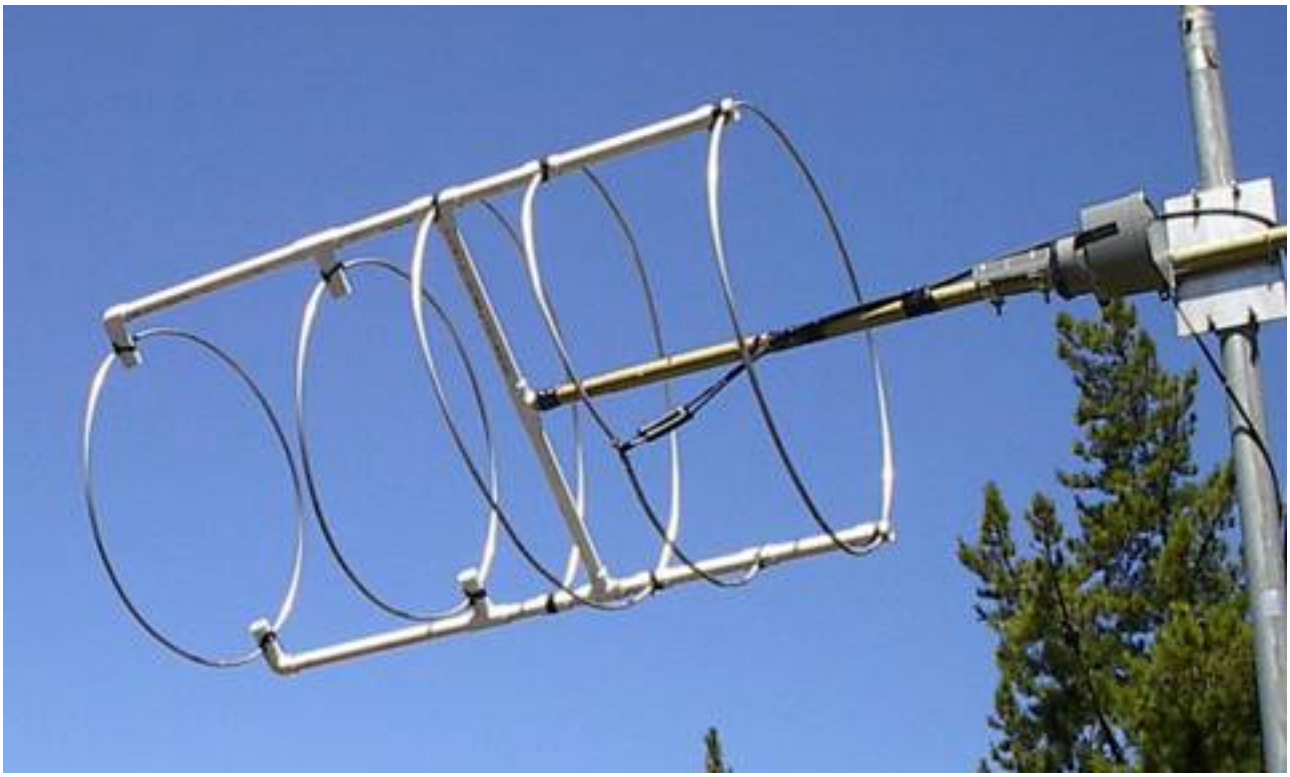
FORMULA: C = 12.5975B - 114.5B^2
C = CORRECTION FACTOR AS A FRACTION OF THE BOOM DIA.
B = BOOM DIA IN WAVELENGTHS
B^2 MEANS B SQUARED

2 METERS
BOOM DIAMETER CORRECTION ADD
2 METERS.
0.750" OR 19.050MM 10.56% .0792" OR 2.02MM
0.875" OR 22.225MM 12.14% .1062" OR 2.70MM
1.000" OR 25.400MM 13.66% .1366" OR 3.47MM
1.125 OR 28.575MM 15.13% .1702" OR 4.32MM
1.250" OR 31.750MM 16.54% .2068" OR 5.25MM
1.375" OR 34.925MM 17.90% .2462 OR 6.25MM
1.500" OR 38.100MM 19.21% .2882" OR 7.32MM
1.750" OR 44.450MM 21.67% .3792 OR 9.63MM
2.000" OR 50.800MM 23.91% .4783" OR 12.15MM
20.000MM 11.04% 2.21MM
28.000MM 14.87% 4.16MM
30.000MM 15.78% 4.73MM
32.000MM 16.66% 5.33MM
38.000MM 19.17% 7.29MM

Insulators can be commercial shoulder insulators and keepers or as simple as

heat shrink tubing with the element held in place by hot melt glue or epoxy glue.

The Quadix



The Quadix is a circularly polarized parasitic array, a hybrid between a helix and a quad (or loop yagi). As is the case with other yagi type parasitic arrays, it has a reflector and as many directors as desired. A two turn helix is used as the driven element. The helix is driven ¼ turn from the end of the helix closest to the reflector. Its impedance is a nearly constant 300 ohms across a wide band for a conductor size of 3/8 inch. As with off center fed antennas, the feed point is not balanced, so a current balun is required. A tv/fm balun works well. Or there are many other balun possibilities (see below). I will now describe a 4 element 2 meter quadix.

Its bandwidth exceeds 144 – 148 MHz. Over the 144 – 148 MHz band, its free space gain is simulated to be about 10.5 dBi, and axial ratio less than 2 dB. The SWR is measured to be less than 1.2 when used with an innovative 300 ohm to 50 ohm balun transformer (described below), and axial ratio measured to be less than approximately 1 dB across the repeater output band 145.2 – 147.4 MHz, the source of my test signals.

The support structure is made from 3/4 inch PVC pipe.

The elements are made of 3/8 inch aluminum refrigerator tubing, obtained locally (Sunnyvale, California) from Orchard Supply Hardware. If aluminum refrigerator tubing is not available, you can use copper refrigerator tubing.

The total length of the helix driven element is 179 3/4 inches. Mark it at 44 15/16 inch intervals to show where it passes through the PVC pipe. Also mark it 22 ½ inches from one end. This will be where the balun is attached. The total length of the reflector will be 89 9/16 inches. The total length of the directors is 76 3/16 inches. Mark each of them at the halfway point.

Before assembling the antenna, temporarily connect the ends of each of the elements together and shape them into circles. The driven element should be shaped into a two turn circle.

Assemble the antenna by feeding the elements through the holes drilled in the PVC assembly. The ends of the directors and reflector can be connected together by flattening and securing with bolts, by slotting one end and squishing it down to fit into the other end and securing with a bolt, or by fitting a sleeve over the two ends and securing with a pipe clamp. If one of the overlap methods is chosen, don't forget to allow for this when cutting the tubing.

Before assembling the helix driven element, cut out a ½ inch section of tubing at the feed point and replace by an insulating material. An insulator fashioned from a piece of PVC coupling works well. For right hand circular polarization, feed the helix clockwise through the holes in the PVC assembly as shown in the picture. Secure the elements to the PVC structure. Connect the antenna to a boom, and the balun and feed line to the antenna, and you are ready to go. If you want to measure the axial ratio against a friend's linear polarized signal (or local repeater signals), you need to attach a small rotator to the boom.

A 300 ohm to 50 ohm quarter wave balun transformer that is simple to make is shown.

It consists of a quarter wave transformer made from RG-58 and RG-59 in series. The impedance of the quarter wave transformer is 53 plus 73 equals 126 ohms, which transforms 300 ohms to 53 ohms. The balun action comes from the fact that the two sides of the 126 ohm transmission line (the center conductors of the coaxes) are isolated from each other. The balun is improved by puting approximately 8 ferrite beads onto each line. It has been pointed out to me by BB that there should be a few beads also on the 50 ohm line near where it connects to the balun.

Ross Anderson W1HBQ July 5, 2002 April 19, 2004

E-mail me at ross_anderson@comcast.net

Notes and References:

Kraus, Antennas, second edition, chapter 7, "The Helical Antenna".
Maxwell, "Some Aspects of the Balun Problem"
http://home.iag.net/~w2du/Reflections2Chapter21SomeAspectsoftheBalunProblem.pdf
Belrose, "The Off-Center-Fed Dipole Revisited: A Broadband, Multiband Antenna"
http://www2.arrl.org/members-only/tis/info/pdf/9008028.pdf
Belrose, "Transforming the Balun", QST, June 1991, pp 30-33.
Rosser, "Use Low-Loss 'Window' Ladder Line for Your 2-Meter Antenna", The ARRL Antenna Compendium, Vol 6, pp 165-167.
The beads I used were FB-43-5622. http://www.amidon-inductive.com/associates_prod_largerbeads.htm

My homepage “Ross’s Antennas” with links to my other pages .
http://home.comcast.net/~ross_anderson

SLIM-JIM 2m ANTENNA

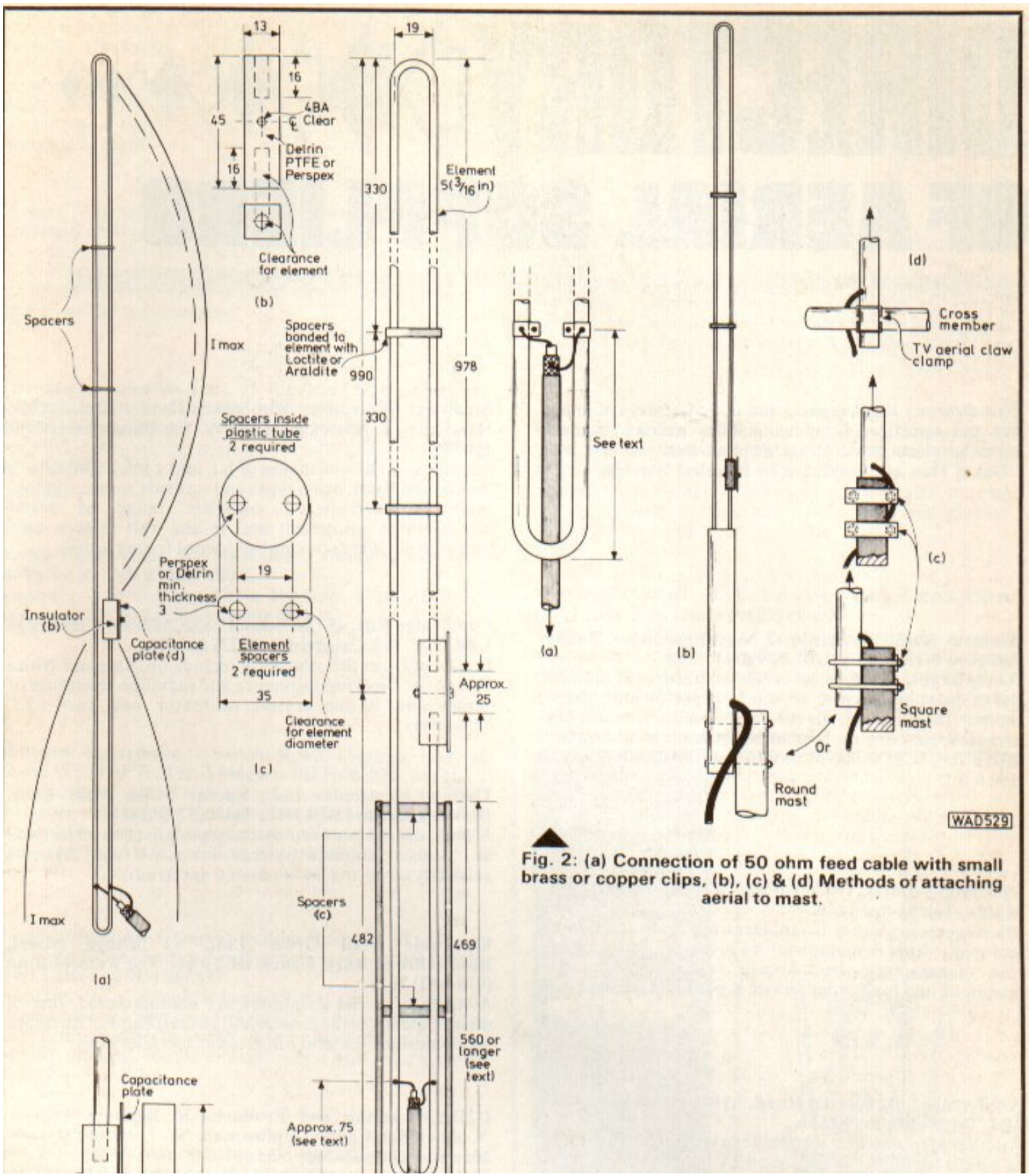
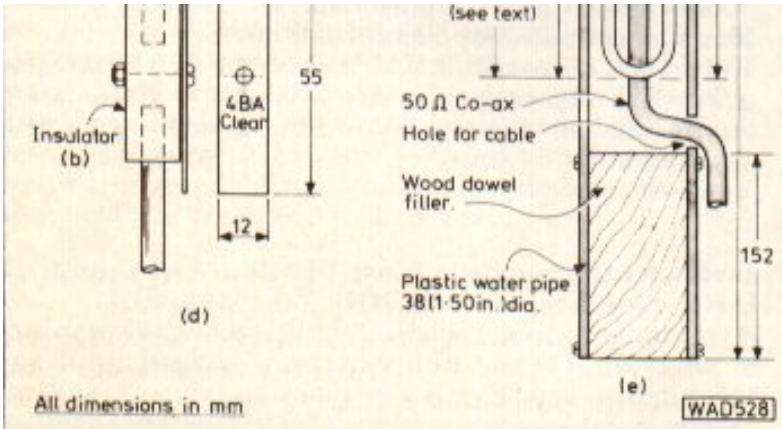


Fig. 2: (a) Connection of 50 ohm feed cable with small brass or copper clips, (b), (c) & (d) Methods of attaching aerial to mast.



All dimensions in mm

WA0528

Fig. 1: (a) General configuration showing voltage and current distribution. (b) Details of capacity plate insulator. (c) Spacer details. (d) Capacity plate details. (e) General view including details of final enclosure in plastics tube.

THE SPERRTOPF ANTENNA FOR THE TWO-METER BAND

By ON4CFC, Pascal Veeckmans ~ Translation by Jef Verborgt

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Cheap Yagi Antennas for VHF/UHF

by Kent Britain, WA5VJB
edited by John Maca, AB5SS

[***Editors notes:*** *The antennas described in this article were built as the result of several discussions between Kent and a Cuban radio operator. While there are plenty of high performance antenna designs, most of the parts required to build them are not available in Cuba. There just isn't an EPO or Radio Shack available in Cuba. Kent accepted this as a challenge to design a really good antenna that could be built with little more than ground wire, coax and a wooden boom. Using the latest antenna design software, he has developed several variations for 144 thru 1296 MHz. Apparently, the designs work very well... Kent entered the 432 MHz version in a recent antenna contest and lost by 0.2 dB to a Midwest ham who had copied his design. Though disappointed in losing, it did prove to Kent that the antennas can be easily replicated with consistant performance.*]

If your planning to build an EME array, don't use these antennas. But, if you want to put together a Rover station with less than $500 in the antennas or just want a good antenna for the home, read on.

These antennas are relatively small, easily constructed from common materials/tools and have surprising performance. The feed method is greatly simpified by directly soldering the coax to the driven element. No baluns or gamma matches are used in this design. This simplified feed uses the structure of the antenna itself for impedance matching. The spacing of the director and reflector elements from the driven element directly affects the feed point impedance of the antenna. So, the design starts with the feed (driven element) and the elements are built around it. Typically, a high gain antenna is designed in the computer, then you try to come up with a matching arrangement for a 31.9 Ohm feed! For the cost about 0.5 dB of gain, these antennas make some design compromises for the feed impedance, use an asymmetrical feed and make trade offs for a very clean pattern. But, they allow simple measurements, have wide bandwidth, the ability to grow with the same element spacing AND... you can build these antennas for $5!!!!

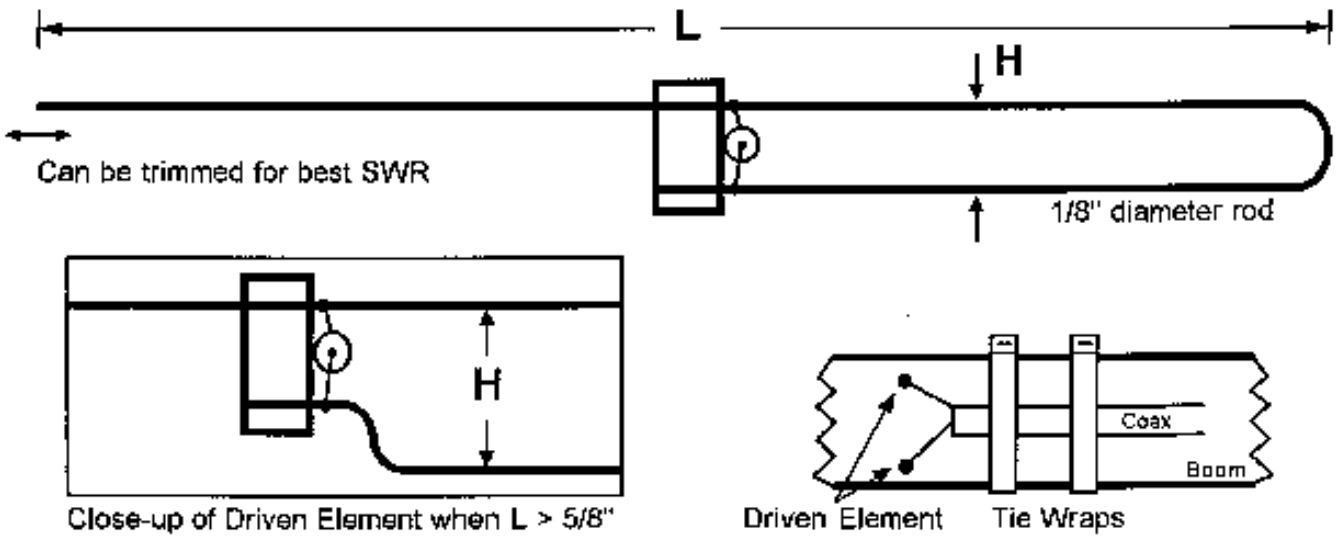
The booms used for these antennas is 1/2" X 3/4" wood. The elements have been made from silicon bronze welding rod, aluminum rod, hobby tubing and solid ground wire with no change in performance. Since you want to be able to solder to the driven element, silicon bronze welding rod, hobby tubing and #10 or #12 solid copper wire have been used and work fine. A drop of "Super Glue", epoxy or RTV is used to hold the elements in place. A good coat of Polyurethane should be applied to the wooden boom to protect it from the weather. A polyurethane varnished 902 MHz version has been in the air for a year now with little deterioration in performance.

And now for the antenna designs. These antennas have been carefully designed to have the highest dB's/Dollar ratio of anything around They were designed with YagiMax, tweaked using NEC and the driven elements experimentally determined on the antenna range. The driven element design is the same for all frequencies except for the length (L) and separation (H). See Figure 1 for details on the driven element. All dimensions are in inches.

SIGs

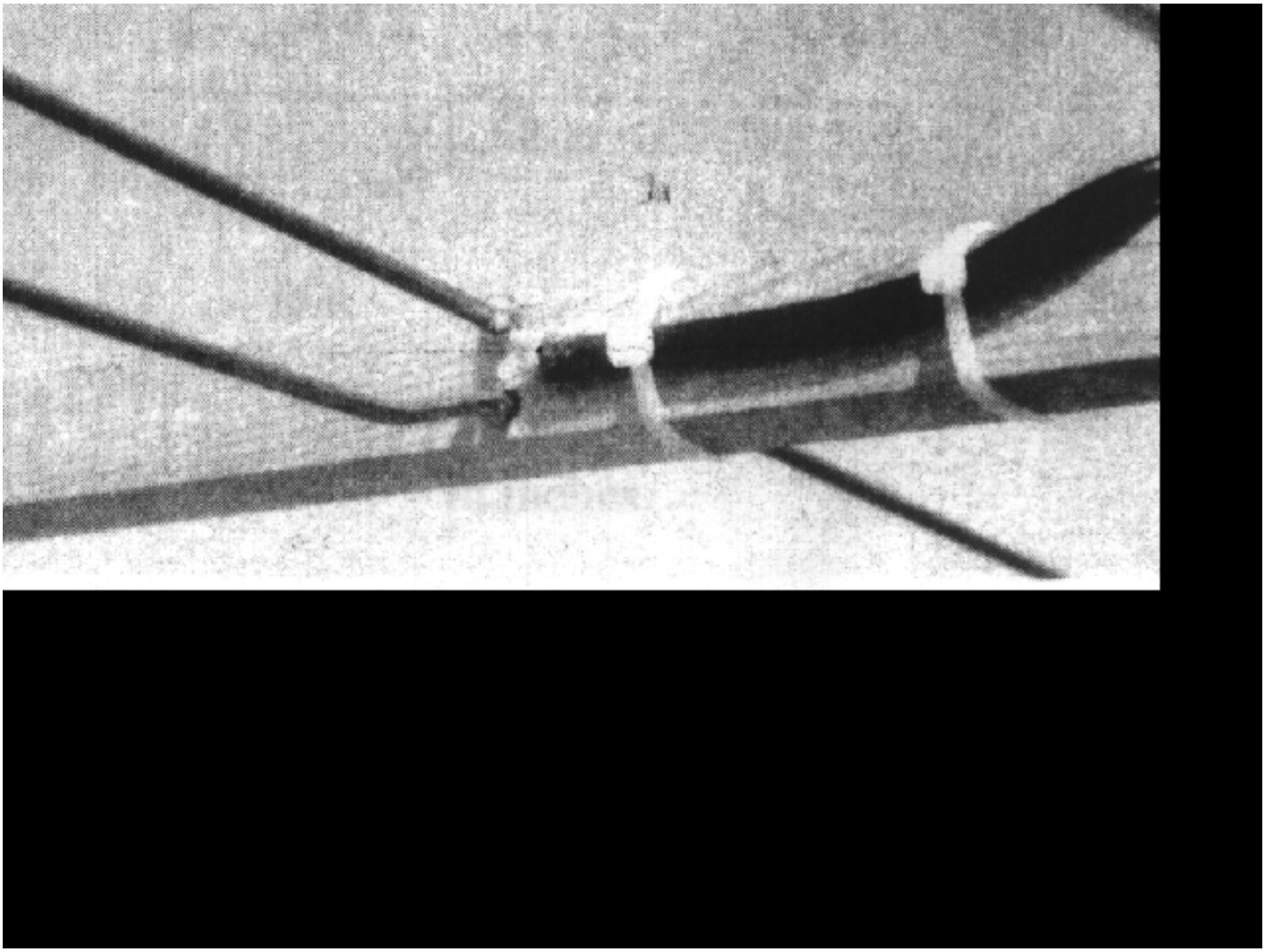
Amateur Links

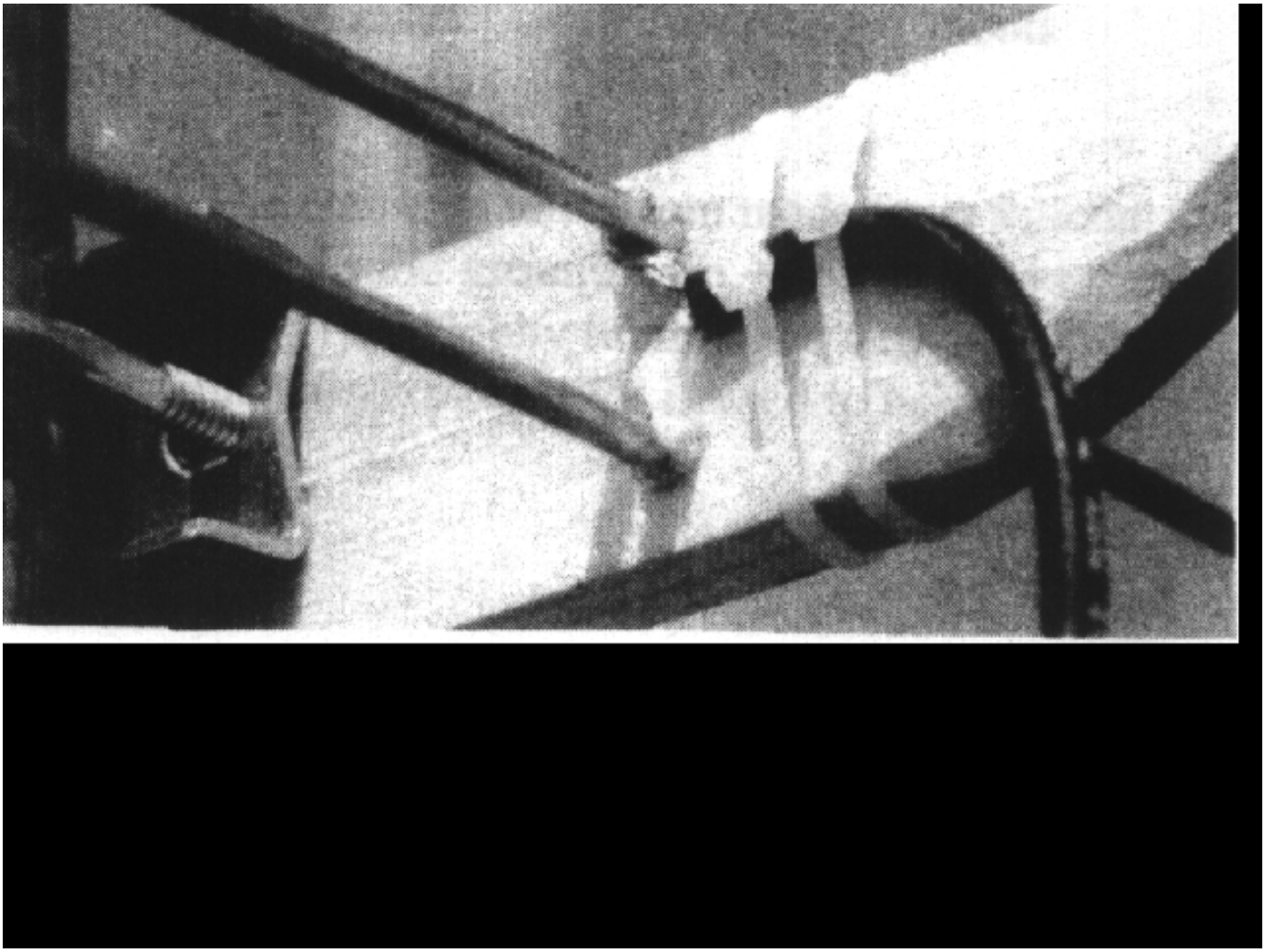
Driven Element Construction (*all versions*)



Close-up of Driven Element when L > 5/8"

Driven Element Tie Wraps





144 MHz. This antenna is peaked for 144.2 MHz but performance is still good at 146.52 (emergency use only!) Driven element dimensions are L = 38.5" and H = 1.0" Elements are 1/8" diameter.

144 MHz		**REF**	**DE**	**D1**	**D2**	**D3**	**D4**
3 Element	Length Spacing	41.00 0.00	8.50	37.00 20.00			
4 Element	Length Spacing	42.00 0.00	8.50	37.50 19.25	33.00 40.50		
6 Element	Length Spacking	40.50 0.00	7.50	37.50 16.50	36.50 34.00	36.50 52.00	32.75 70.00

222 MHz. This antenna is peaked for 222.1 MHz but performance bearly changes at 223.5 MHz. Driven element dimensions are L = 24.5" and H = 1.0" Elements are 3/16" diameter.

222 MHz		**REF**	**DE**	**D1**	**D2**	**D3**	**D4**
3 Element	Length Spacing	26.00 0.00	5.50	23.75 13.50			

4 Element	Length Spacing	26.25 0.00	5.00	24.10 11.75	22.00 23.50		
6 Element	Length Spacing	26.25 0.00	5.00	24.10 10.75	23.50 22.00	23.50 33.75	21.00 45.50

432 MHz. This antenna is peaked for 432.1 MHz. At this frequency, this antenna is getting very practical and easy to build. Driven element dimensions are L = 13.0" and H = 3/8" Elements are 1/8" diameter.

432MHz		**REF**	**DE**	**D1**	**D2**	**D3**	**D4**	**D5**	**D6**	**D7**	**D8**	**D9**
6 Element	Length Spacing	13.50 0.00	2.50	12.50 5.50	12.00 11.25	12.00 17.50	11.00 24.00					
8 Element	Length Spacing	13.50 0.00	2.50	12.50 5.50	12.00 11.25	12.00 17.50	11.00 24.00	12.00 30.75	11.25 38.00			
11 Element	Length Spacing	13.50 0.00	2.50	12.50 5.50	12.00 11.25	12.00 17.50	12.00 24.00	12.00 30.75	12.00 38.00	11.75 45.50	11.75 53.00	11.00 59.50

902/903 MHz. This was the first antenna I built using the antenna to control the driven element impedance. The 2 1/2' length has proven practical, so I haven't built any other versions. Driven element dimensions are L = 5.7" and H = 1/2" Elements are 1/8" diameter.

902/903 MHz		**REF**	**DE**	**D1**	**D2**	**D3**	**D4**	**D5**	**D6**	**D7**	**D8**
10 Element	Length Spacing	6.20 0.00	2.40	5.60 3.90	5.50 5.80	5.50 9.00	5.40 12.40	5.30 17.40	5.20 22.40	5.10 27.60	5.10 33.00

1296 MHz. This antenna is the veteran of several "Grid Peditions" but I have yet to actually measure the gain. Dimensions must be followed with great care. The driven element is small enough to allow 0.141 semi-rigid coax to be used instead of RG-58. Silicon Bronze welding rod was used for the elements but any material can be used. Driven element dimensions are L = 4.0" and H = 1/2" Elements are 1/8" diameter.

1296 MHz		**REF**	**DE**	**D1**	**D2**	**D3**	**D4**	**D5**	**D6**	**D7**	**D8**
10 Element	Length Spacing	4.30 0.00	1.70	3.90 2.80	3.80 4.00	3.75 6.40	3.75 8.70	3.65 12.20	3.60 15.60	3.60 19.30	3.50 23.00

OTHER VERSIONS

421.25 MHz ATV. 421 MHz Vestigial Sideband video is popular in North Texas for receiving the FM video repeaters. The driven element for these antennas is designed for an impedance of 75 ohms. So RG-59, or an `F' adapter to RG-6, can be directly connected to a cable TV converter/Cable Ready TV on channel 57. Driven element dimensions are L = 13.0" and H = 1/2" Elements are 1/8" diameter.

Spacing is the same for all versions.

421 MHz ATV		REF	DE	D1	D2	D3	D4	D5	D6	D7	D8	D9
6 Element	Length	14.00		12.50	12.25	12.25	11.00					
8 Element	Length	14.00		12.50	12.25	12.25	12.00	12.00	11.25			
11 Element	Length Spacing	14.00 0.00	3.00	12.50 6.50	12.25 12.25	12.25 17.75	12.00 24.50	12.00 30.50	12.00 36.00	11.75 43.00	11.75 50.25	11.50 57.25

450 MHz FM. Yea, I understand it's FM, but sometimes a newcomber needs a cheap antenna to get into a repeater or give you a simplex QSO during a contest. Driven element dimensions are L = 12.0" and H = 3/8" Elements are 1/8" diameter. Spacing is the same for all versions.

450 MHz FM		REF	DE	D1	D2	D3	D4
6 Element	Length Spacing	13.00 0.00	2.50	12.10 5.50	11.75 11.00	11.75 18.00	10.75 28.50

435 MHz AMSAT. The larger versions have not been fully tested and I appreciate the help and motivation from KA9LNV for these antennas. Updates and performance evaluations are planned for a later edition of the AMSAT Journal. A high Front-to-Back ratio was the major design consideration for all versions. The computer predicts 30 dB F/B for the 6 element and over 40 dB for the others. NEC predicts 11.2, 12.6, 13.5 and 13.8 dBi for the 6, 8, 10 and 11 element respectively. Using 3/4" square wood makes it easy to build two antennas on the same boom for cross- polarized operation. Offset the two antennas 6 1/2" and feed in phase for Circular Polarization. Or, just build one antenna for portable operation. Driven element dimensions are L = 13.0" and H = 1/2" Elements are 1/8" diameter. Spacing is the same for all versions.

435 MHz AMSAT		REF	DE	D1	D2	D3	D4	D5	D6	D7	D8	D9
6 Element	Length	13.40		12.40	12.00	12.00	11.00					
8 Element	Length	13.40		12.40	12.00	12.00	12.00	12.00	11.10			
10 Element	Length	13.40		12.40	12.00	12.00	12.00	12.00	11.75	11.75	11.10	
11 Element	Length Spacing	13.40 0.00	2.50	12.40 5.50	12.00 11.25	12.00 17.50	12.00 24.00	12.00 30.50	11.75 37.75	11.75 45.00	11.75 52.00	11.10 59.50

7-Element-Yagi for the 2m-Band with the 28-Ohm-DK7ZB-Match

This antenna was described in the issue 2/2000 of the magazine "FUNKAMATEUR" , report "Kurze Yagis für das 2m-Band in bewährter 28-Ohm-Technik".

Short data: 11dBd gain, boomlength 3,30m, very good pattern. For the match see: **DK7ZB-Match**

At the webpage "DOWNLOAD" you can load the report as PDF-File (german language!).

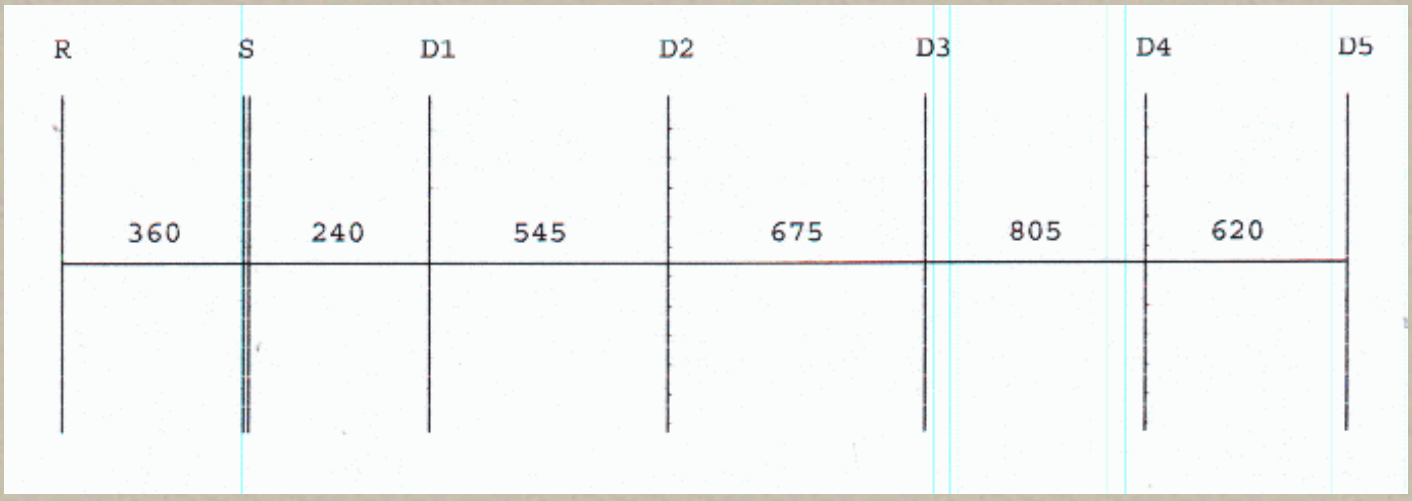
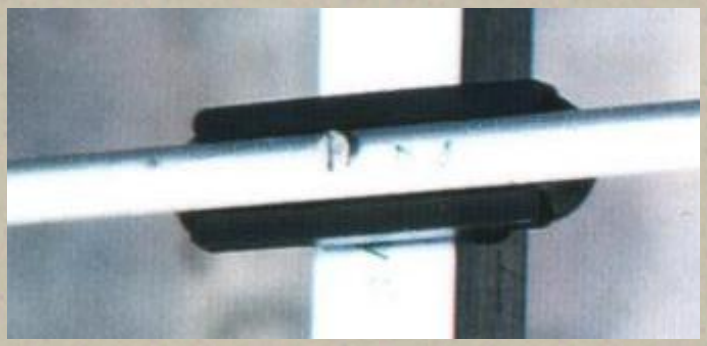


Table of Elements with 4-10mm Diameter

The radiator has always a diameter of 10mm!

El.-Diameter	4mm	6mm	8mm	10mm
Reflector	1019mm	1016mm	1014mm	1011mm
Radiator	978mm	978mm	978mm	978mm
Director 1	959mm	951mm	945mm	940mm
Director 2	933mm	924mm	916mm	909mm
Director 3	916mm	906mm	898mm	891mm
Director 4	915mm	905mm	897mm	890mm
Director 5	904mm	894mm	885mm	877mm

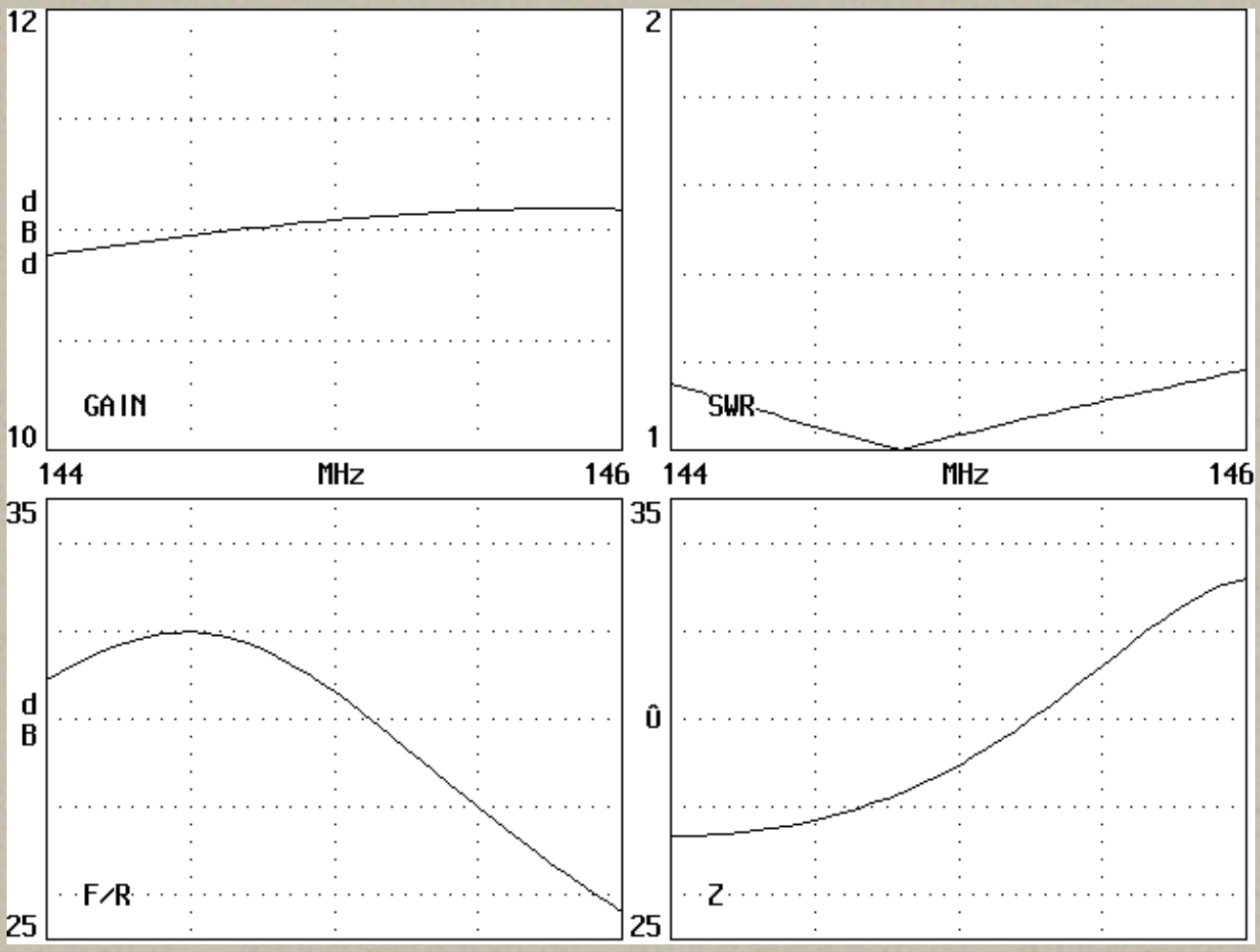


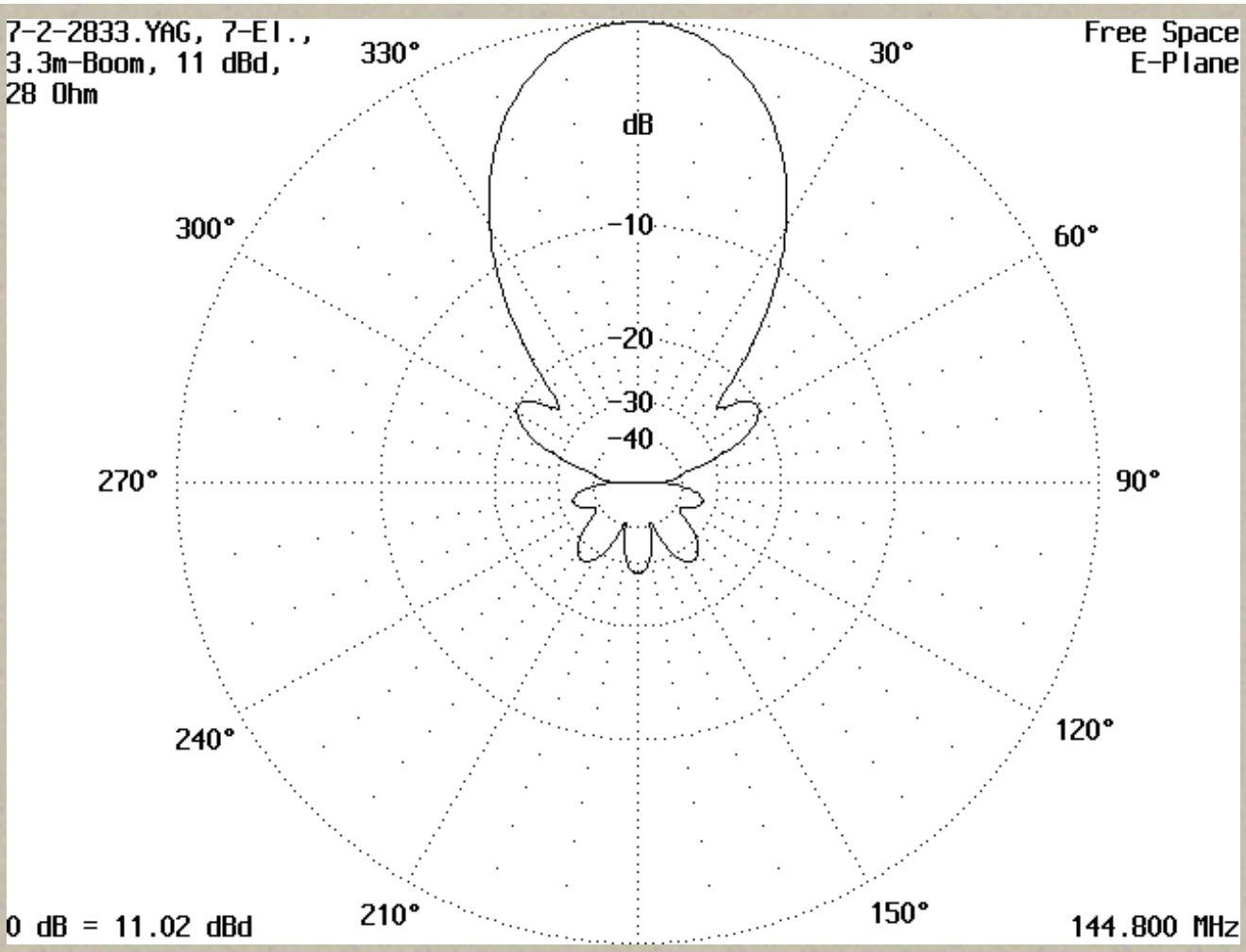
Elements are mounted isolated above the boom with polyamid clamps for TV-Yagis (Konni)

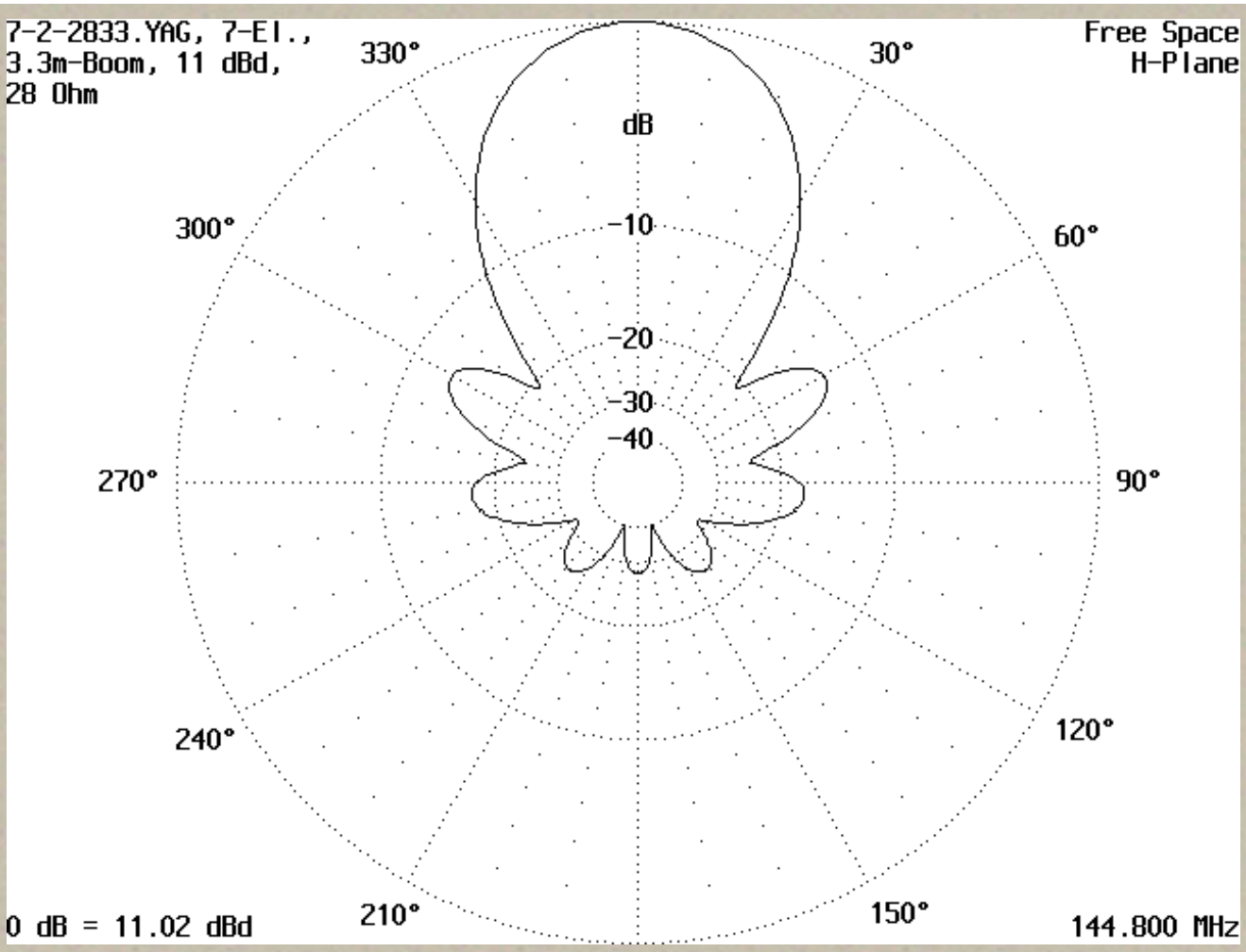
The screw in the center of the element has no influence to the length of the elements!

Gain, SWR, Front/Rear and Feedpoint-Impedance

Best F/R in the CW/SSB-part of the 2m-Band




7-2-2833.YAG, 7-El.,
3.3m-Boom, 11 dBd,
28 Ohm
Free Space
E-Plane
dB
-10
-20
-30
-40
330°
30°
300°
60°
270°
90°
240°
120°
210°
150°
0 dB = 11.02 dBd
144.800 MHz



Further details and pictures can be found on the homepages of

DC6GF

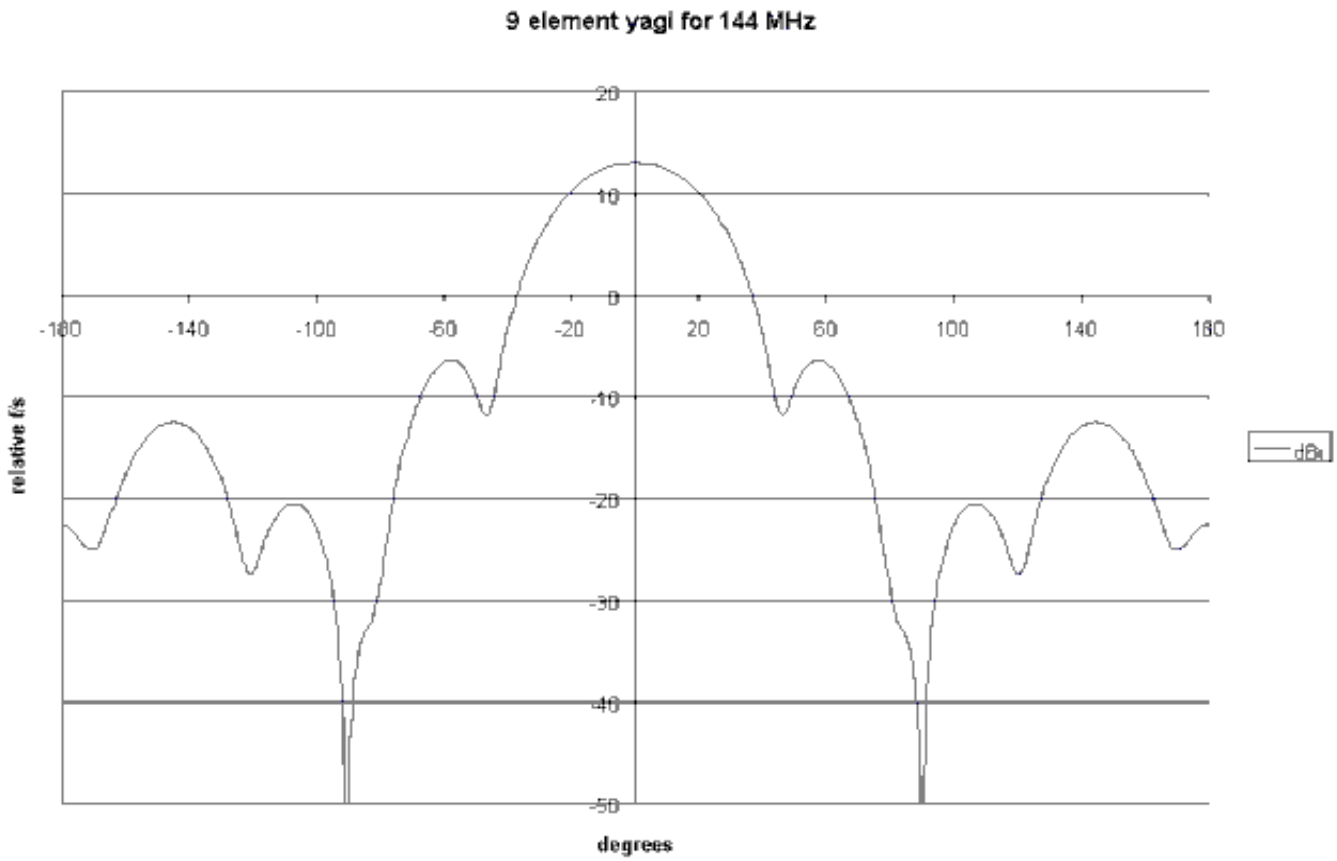
DF8GH

This page is still under construction, but I have more to report on this yagi.

This yagi is based on a design by JM1SZY for 50 MHz that he developed using YO. On 6 Metres, JM1SZY designed this antenna for good gain, fairly wide bandwidth, and a great F/B ratio.

When I scaled it to 2 Metres, it had another quality, it was just a shade over 11 feet in length! I don't know about the rest of the world, but in North America, wood and aluminum tubing is commonly sold in 12 foot lengths. So this antenna could be built with either a single length of 12 foot AL tube, or, if you can find .058" wall aluminum, out of 4 foot sections of tubing. Waste not, want not!

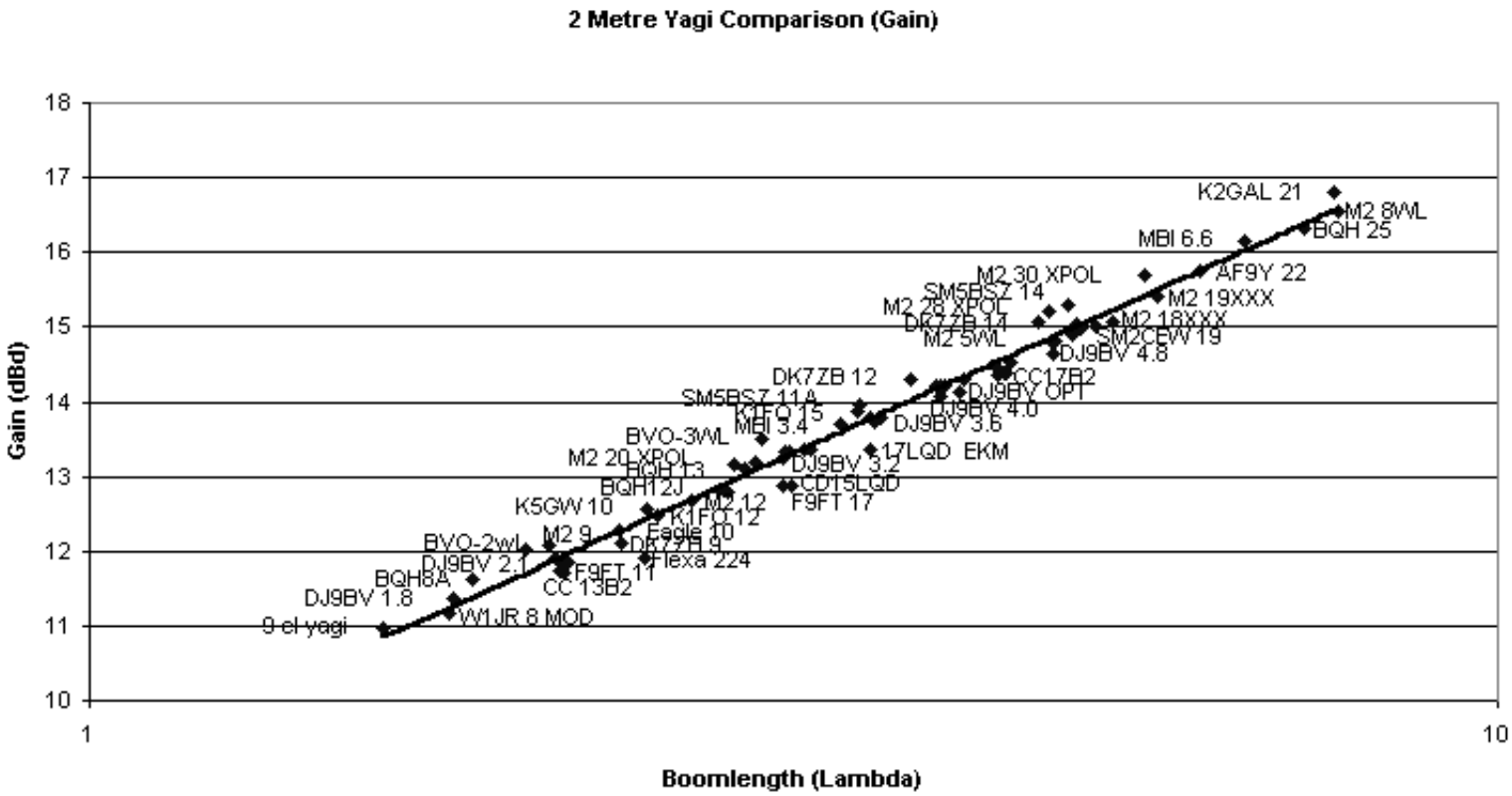
Here's it's free-space pattern on 144.2. It's gain is in line with the DL6WU formula for gain vs. boomlength. It's front-to-back ratio is quite good, IMHO. The far-field data came from NEC2 and was plotted using MS-Excel.



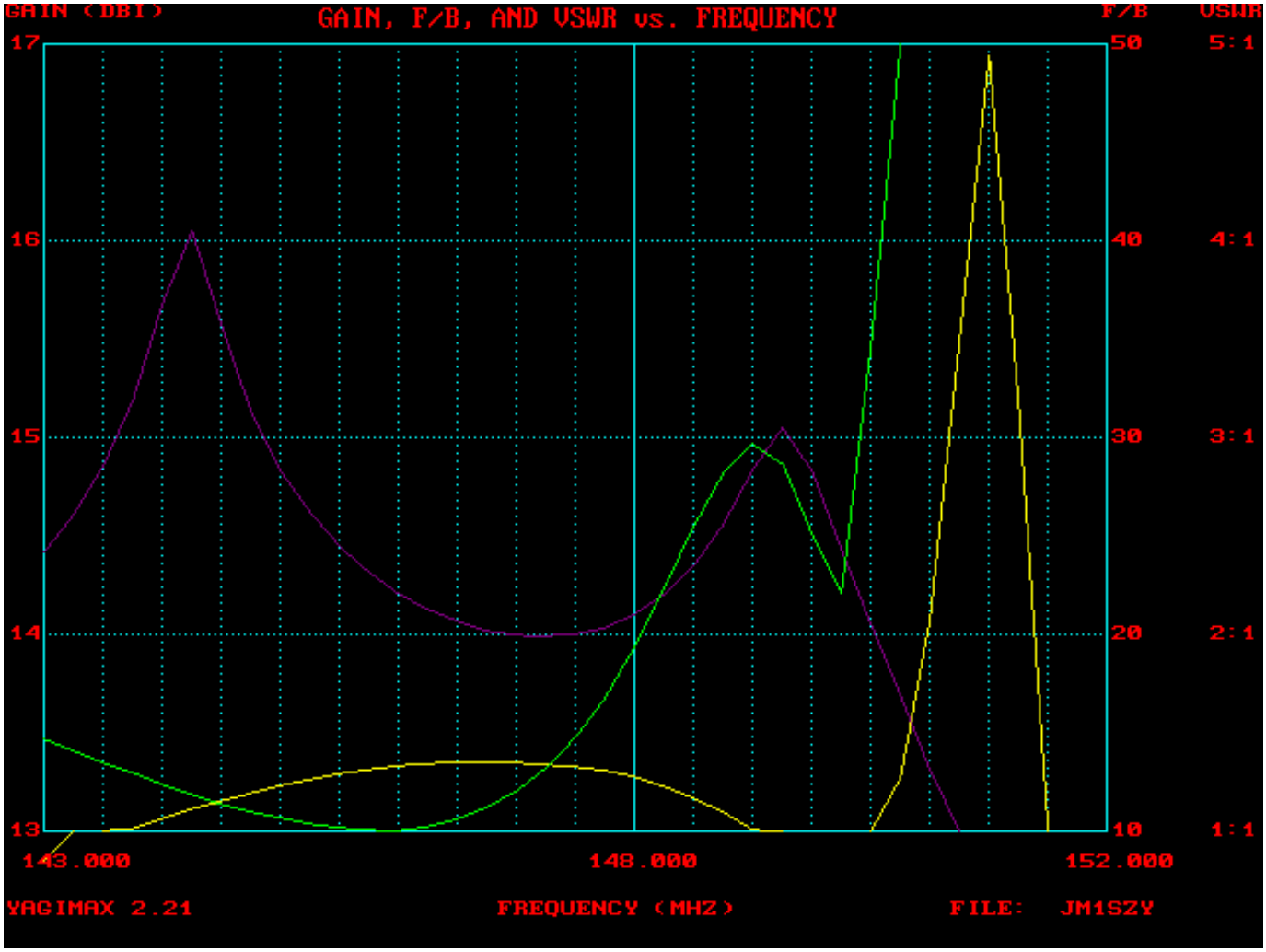
I decided have the gain peak higher up in the band from 144.2. According to K1FO, antennas designed this way are better behaved in a stacked array. I also hope to use it for OSCAR work, so having good gain near 145.9 MHz was a desirable feature, as well.

VE7BQH has done considerable work comparing yagis for 2 Metres. Again, using MS-Excel, I've taken the BQH data, and plotted a trend line showing calculated gain for various antennas. The boomlength axis is logarithmic in scale. Keep in mind, most of these big boys are used for moonbounce, so you'll find the 9 element antenna plotted at the bottom left corner: it is considerably shorter (and subsequently of considerably less gain) than the examples in BQH's table. The data used in this chart can be found at:

http://web.wt.net/~w5un/144gt1.htm

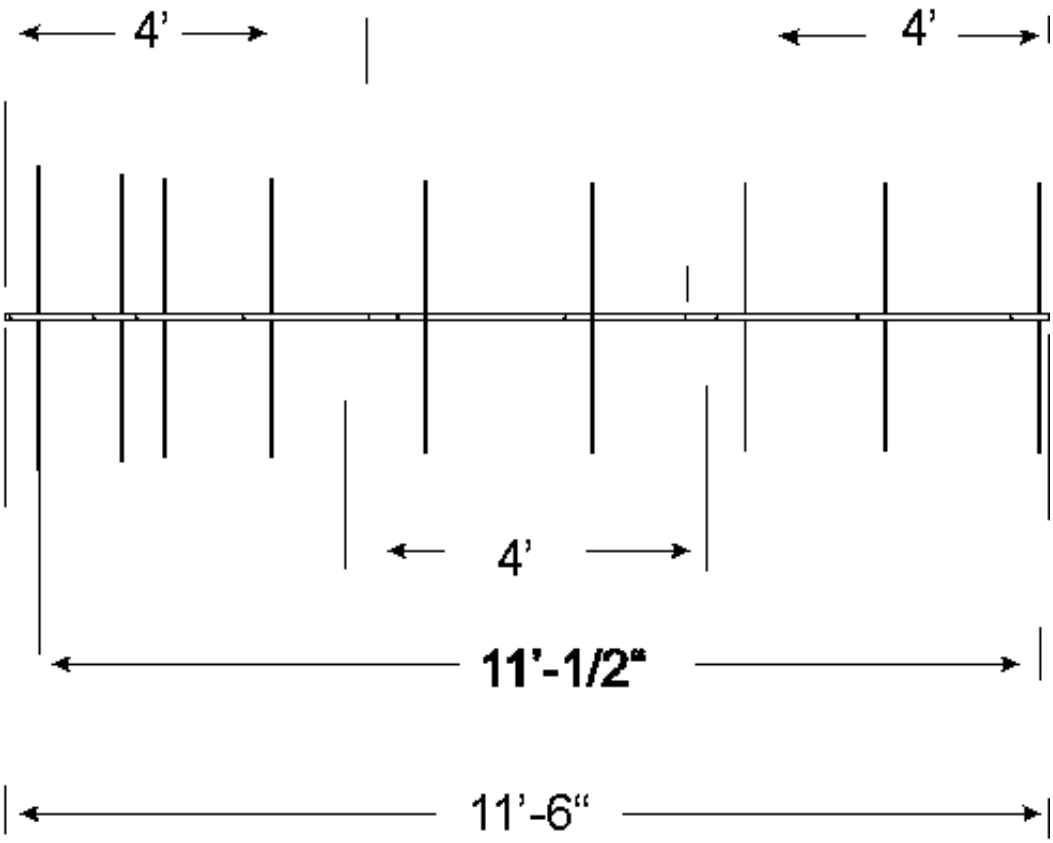


Here's a graph of gain (in dBi), F/B ratio and VSWR vs frequency that was plotted by YagiMax. I'm using a T-match with a half-wave coaxial balun, and the SWR bandwidth is similar to that shown in the graph. The gain is shown in yellow, the F/B ratio in magenta, and the VSWR in green.



Free Space Dimensions

Element	X axis (MM)	Y axis (MM)	Diameter (MM)
REF	0.000	513.584	4.76
DE	279.128	499.083	4.76
D1	424.349	478.218	4.76
D2	783.058	467.091	4.76
D3	1301.530	459.510	4.76
D4	1865.040	453.285	4.76
D5	2376.020	442.401	4.76
D6	2848.920	452.346	4.76
D7	3365.820	445.000	4.76


4'
4'
4'
11'-1/2"
11'-6"

Simpleton's Guide to: Quickie Antennas

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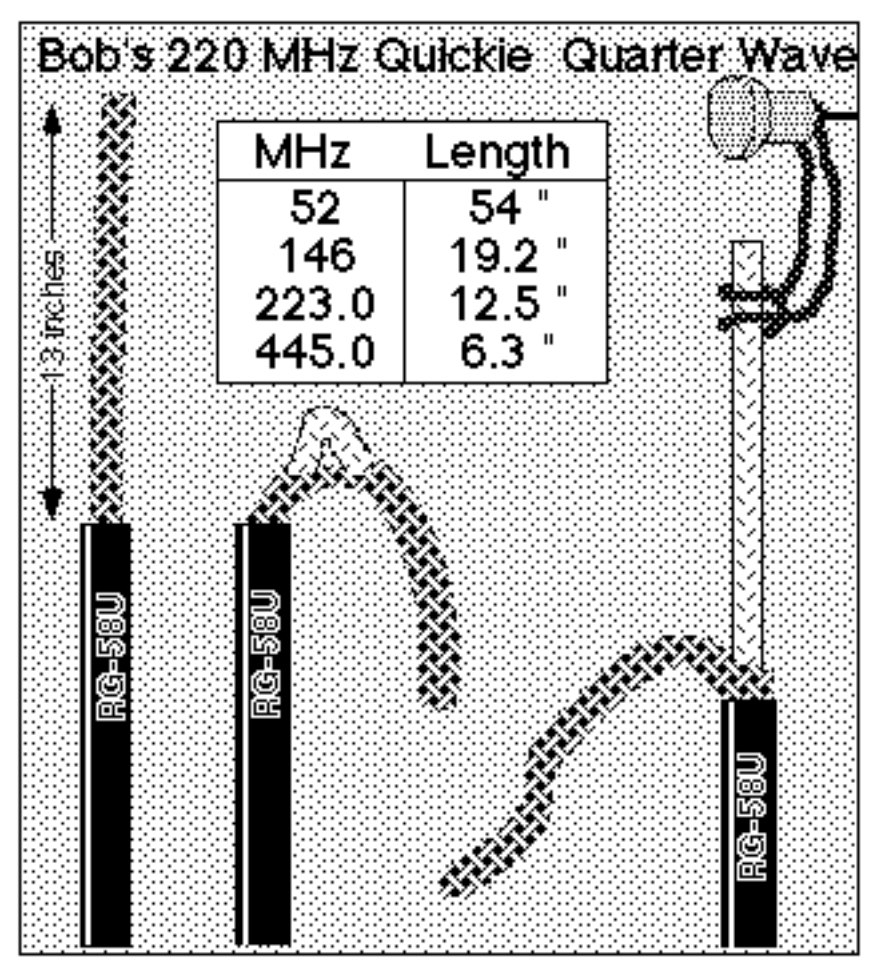
Rebuilding a ham shack after moving into a new home is a very big job. Even if you only have a single VHF radio, running the coax and planting a new antenna system on the roof seems to take more effort each time you do it. Once you get used to drilling holes all the way through a wall for the coax, you are faced with how to support the antenna mast on the roof. If you are only renting the home, you may be tempted to strap the mast on the side of the house and limit the height to just over roof level.

After moving my family into a new home a few months ago, I set aside a weekend to work on my ham shack and antenna system. In my old house, I had an antenna for each band, 2 meters, 220 MHz and 440MHz. Since we are only renting this new home, I wanted to make use of a new multiband antenna. I drilled the holes, built the antenna and mast and bolted it to the side of the house. A triplexer (three port RF splitter) would breakout each band for connection to my three radios. I have used duplexers before, but this would be my first experience with a triplexer.

I was ready to turn the radio on and test my new system. The radios were receiving properly, but my transmitter signal was not making connections to the local repeaters. I connected the antenna to each radio and everything was fine. I connected the triplexer and hooked up a SWR meter and discovered that the SWR through the triplexer was over 5 to 1. I couldn't understand how the receive signals made it through the triplexer but my transmitter could not.

Luckily I had an old duplexer and hooked it up with great success. My 2 meter and 440 MHz radios could now operate properly. The 220 MHz radio was left without an antenna. It was now getting dark, I didn't want to drill another hole for another run of coax and there was no room on the mast for another antenna. I was not going to be defeated by that

malfunctioning triplexer.



I remembered that a 220 MHz quarter wave antenna could be made using a 12 1/2" wire for a radiator and a 13" wire for the ground plane. It was then that the inspiration for a quickie antenna hit me. I had a left over 6 foot piece of coax with a PL connector on one end. I removed 13" of outer insulation and exposed the shield. I measured down 12 1/2 inches and forced open the shield and folded the coax over and pulled out the center conductor with insulation through the hole in the shield.

I now had a 220 MHz quarterwave antenna!! I hung it from the ceiling in my shack with a rubberband and a tack. I hooked it up to the meter and measured the SWR. It was perfect.

This antenna can be built for any frequency you may need.

Just make sure the length of the radiator is correct. If you need to look up the exact length for each band, see the Simpleton's Guide to Quarterwave Antennas. You may attach the radiator to a wood board to help support it.

You must use RG-58 or similar size coax. The larger RG-8 style is much too hard to work with when you attempt to pull out the center insulation. You may be tempted to unbraid the shield instead of opening a hole and pulling out the center insulation. Try unbraiding the wire and you'll get so frustrated with the braid tangling in themselves, you'll cut off the work and start over.

Good luck and remember to have fun when you work on the hobby we call Amateur Radio.

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J Pole Antenna

The N6ZAV J-Pole makes a great (but not so hidden) transmitter antenna.

Using the above table, cut the tubing to the size shown for the desired band. The only section where size is not critical is the mast section. Use any size you desire for mounting purposes.

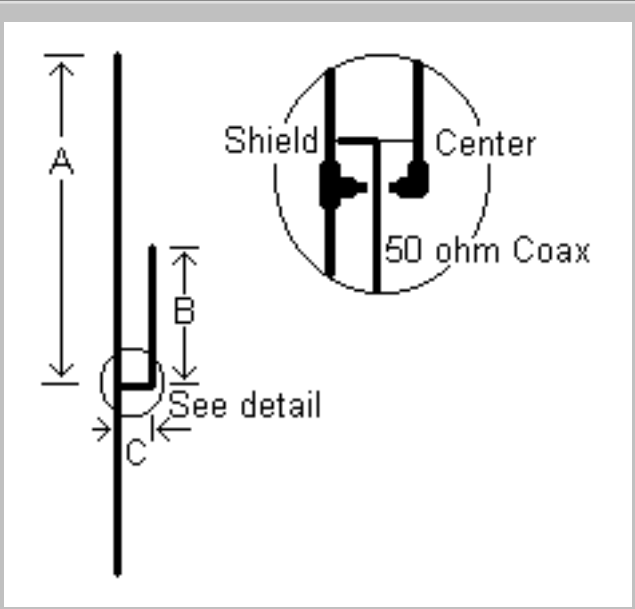
Use steel wool to clean the copper and solder the antenna together as you would with normal copper tubing. Pay close attention to keep the joints straight and snug while doing this. Soldering this together takes quite a bit of heat and a propane torch is the tool of choice.

After the antenna has had a chance to cool, you can attach the coax. Strip the end of the coax that you are going to connect to the antenna and make sure that there is just enough to span the distance between the elements. The cable should be centered between the elements. Strip the center conductor and tin it with solder if it is multi-stranded. Do the same for the braided shield. Using the stainless steel hose clamps, attach the center conductor to the "B" side of the antenna and the shield to the "A" side. Start with the connections about 3" to 5" above the bottom of the "J" and tune with an SWR meter. That's about it! See you on-the-air!

73 de Marty, N6ZAV

```
Parts List:
 1 10' length of 1/2" copper
   tubing schedule "L" or "M"
 1 1/2" copper tubing "T" fitting
 1 1/2" copper tubing elbow
   fitting
 2 1/2" copper end caps
 2 small stainless steel hose
   clamps
 1 50 ohm coax cable with
   connector for radio.

Element Size Table:
```



	2 Meters		220 MHz	
	Total length	Cut Tubing	Total length	Cut Tubing
A	60 5/8"	60 1/4"	39 3/4"	39 3/8"
B	20 1/4"	19 7/8"	13 1/4"	12 7/8"
C	3"	2 1/4"	2"	1 1/4"

This page by Byon Garrabrant N6BG byon@mail.com 12/24/97

Page one

How to build a high gain vertical antenna for the UHF amateur or CB bands

You can use low-cost coaxial cable to make a simple, high performance, omnidirectional vertical antenna that is ideal for both home station and portable applications.

THE collinear antenna has been around a long time. Various versions enjoyed popularity on the amateur VHF and UHF bands in the eras before and after World War II. But the collinear fell out of favour when the Yagi array became popular since the late '50s. The Yagi's popularity is attributable to its feature of giving the 'best bang for the buck'. But it is a beam which requires rotating.

With the rise in popularity of FM operation on the VHF and UHF bands since the '70s, the proliferation of commercial amateur rigs, and the development of repeater networks around the country, the demand for omnidirectional antennas grew apace. A lot of FM activity is mobile, with a degree of base or home station operation, too. For the latter application, an omnidirectional antenna with gain offers distinct advantages, particularly where comparatively low-powered mobile rigs are used at home.

The growth of UHF CB has followed a similar path, boosted by the availability of locally-manufactured transceivers selling alongside imports. Open access repeaters helped the growth of UHF CB, too.

A home-constructed antenna can save you big bucks. Many constructors make up a simple *groundplane* or *coaxial dipole*, which have the advantage of simplicity. However, something that offers a respectable amount of gain and can be assembled with little more effort is a bonus.

The collinear antenna to be described offers considerable gain and improved bandwidth over the conventional groundplane, coaxial dipole, 'Slim Jims' or similar antennas. It is simple to construct and erect since it does not require tuning or pruning, and uses cheap, commonly available 'quarter-inch' RG58 coax.

The word *collinear* means 'in line', the elements of the collinear antenna being placed in line, end to end. Two half wave dipoles placed end to end and fed out of phase make the simplest two-element collinear.

A collinear from coax

To make a collinear antenna from coaxial cable, a number of elements, each an *electrical* half wavelength long, are joined together with the inner conductor and the shield braid transposed at each joint, as illustrated in Figure 1. An even number of elements is required. By transposing the coax's inner and outer conductors at each join, each half wave element is fed out of phase.

I first ran across this form of the collinear in a scientific publication in the early '70s. The published paper described a *monstrous*, 400 metre long, 104 element array used for a

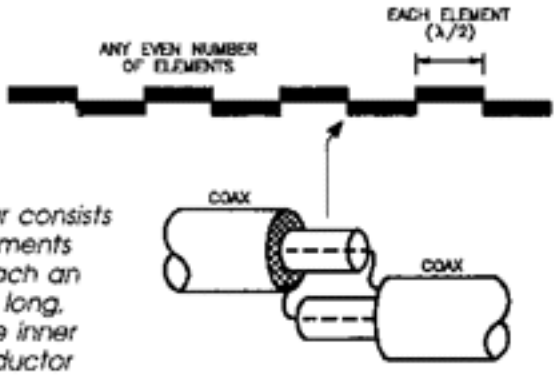


Figure 1. A coaxial collinear consists of an even number of elements made of coaxial cable, each an electrical half wavelength long, joined end to end with the inner conductor and outer conductor transposed at each joint.

advantage of simplicity. However,

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Page three

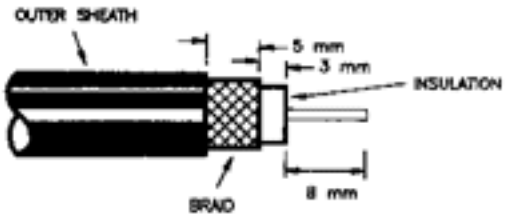

Figure 4. Each element is cut from a piece of RG58 50 ohm coaxial cable and the ends prepared like this.

to any non-conducting support. Dowelling rod from your local hardware store is great for this job and it comes in standard two metre lengths, which is just right. Choose 12.7mm or 19mm diameter dowel to suit yourself.

Now, go through the following procedure step by step and you'll find your collinear goes together quite easily.

1) The very first thing to do is prepare the collinear's support, using a 12.7mm or 19mm diameter wooden dowel rod. This is cheap, readily available and strong enough for the job. The dowel should be thoroughly sealed with an outdoor wood stain or linseed oil, paying particular attention to the ends. Stand it aside to dry properly.

2) Now for the collinear itself. The 'working' length of each element is the distance between the ends of the *braid*. To simplify matters, and to allow for the odd error, cut eight lengths of RG58, each 250mm long for the 70cm amateur band, or 230mm long for the UHF CB band. These lengths make allowance for cutting and stripping back the ends of the elements to make the joints.

3) Prepare each end of *seven* elements, and only one end of the eighth element, as detailed in Figure 4. The eighth element will become the 'top' element of the antenna.

Cut the coax's outer sheath 16mm back from the end using a blunt penknife or hobby knife. It should be blunt so as to avoid nicking the shield braid here. *Do not* unravel the shield braid.

4) Now cut the braid, this time using a sharp knife, 8mm back from the end. Take care not to cut through the dielectric to the centre conductor. Combined use of a sharp knife and sharp, pointed sidecutters can be effective and result in a neat cut.

for a few seconds to heat it, then apply the solder. Use thin gauge, resin-cored solder. But remember to only apply enough solder to lightly 'wet' the conductors.

7) Now to solder the elements together. First slip a 35-40mm length of 9.5mm or 12.7mm diameter heatshrink on each element. Solder the elements together, end to end, as shown in Figure 5. With each joint, after it has cooled, apply silicone sealant to the area of the joint to seal it, then, while the silicone is still plastic, slip the heatshrink tubing over the joint and apply a blast of hot air (hair dryers are great for this). But don't overdo the hot air, though, or you're likely to soften the outer sheath of the coax and possibly damage it.

8) The top element needs to be sealed. Apply a dob of silicone sealant to it, slip on a 30-40mm length of 6.4mm heatshrink while it's still soft, then apply a short blast of hot air to shrink it in place.

9) The next step is to attach what you've just completed to its support. Plastic zip-up cable ties are great for this, as are the plastic zip-lock ties that come with packets of garbage bags. Tie the collinear to the dowel, starting with the top element, putting a tie either side of each joint. The top element should be tied about 50mm below the top end. The other elements should be near the joints.

While the collinear should be laid straight when tying it to the dowel, don't

mechanical support, its prime purpose is protection.

10) Now for the feedpoint and groundplane. You'll have a short length of RG58 left over. Attach a suitable in-line connector, such as a BNC male, to one end and prepare the other end as per Figure 4.

Measure back along the cable, from the end of the shield braid, a quarter wavelength (this time, 'free space' wavelength). For the 70cm amateur band, this is 170mm; for the UHF CB band, 155mm. Mark this point.

Using a blunt knife, or carefully using a sharp knife, make two cuts around the cable's outer sheath, each a few millimetres either side of this point. Slit the sheath between the two cuts and remove the section to expose the shield braid. Using a hot iron, quickly and lightly tin the braid. Slip two 30-40mm lengths of 6.4mm diameter heatshrink down the cable, placing them either side of the exposed shield braid.

11) Cut two lengths of tinned copper wire or brazing rod to size: each 360mm long for the 70cm amateur band, and 330mm long for the UHF CB band (see Figure 3). If you're using tinned copper wire, straighten it first. This can be done by clamping one end in a vise, grasping the other end with a pair of pliers and giving it a good tug. It will bow a bit after you take it out of the vise, but then you can straighten it easily by hand. Tin the centre of each groundplane element.

12) Now attach the prepared cable to

Combined use of a sharp knife and sharp, pointed sidecutters can be effective and result in a neat cut.

5) Next, cut back the dielectric 8mm back from the end to expose the centre conductor. Do this carefully so you don't nick the stranded centre conductor wires. Otherwise, later you may get a break in the centre conductor, or a stray strand may short the joint. Either way, your antenna won't work properly.

6) With the ends of all the elements prepared as per Figure 4, now tin the exposed centre conductor and shield braid on each. Use a hot iron, preferably a temperature controlled type. A flat-faced ('spade') tip is best for this job. Apply the tip to the part to be tinned near the joints.

While the collinear should be laid straight when tying it to the dowel, don't apply too much tension to avoid fracturing the soldering at the joints. Don't depend on the heatshrink for

can straighten it easily by hand. Tin the centre of each groundplane element.

12) Now attach the prepared cable to the feedpoint, making a joint as per Figure 5. Seal it and cover it with heatshrink. Put a tie either side of the

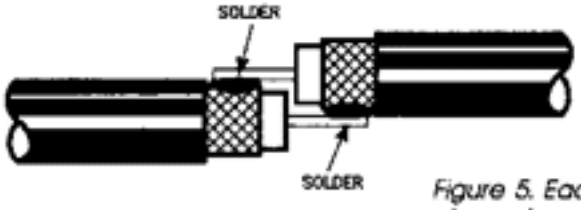

Figure 5. Each joint between the elements, and the joint at the feedpoint, is made like this.

Page four

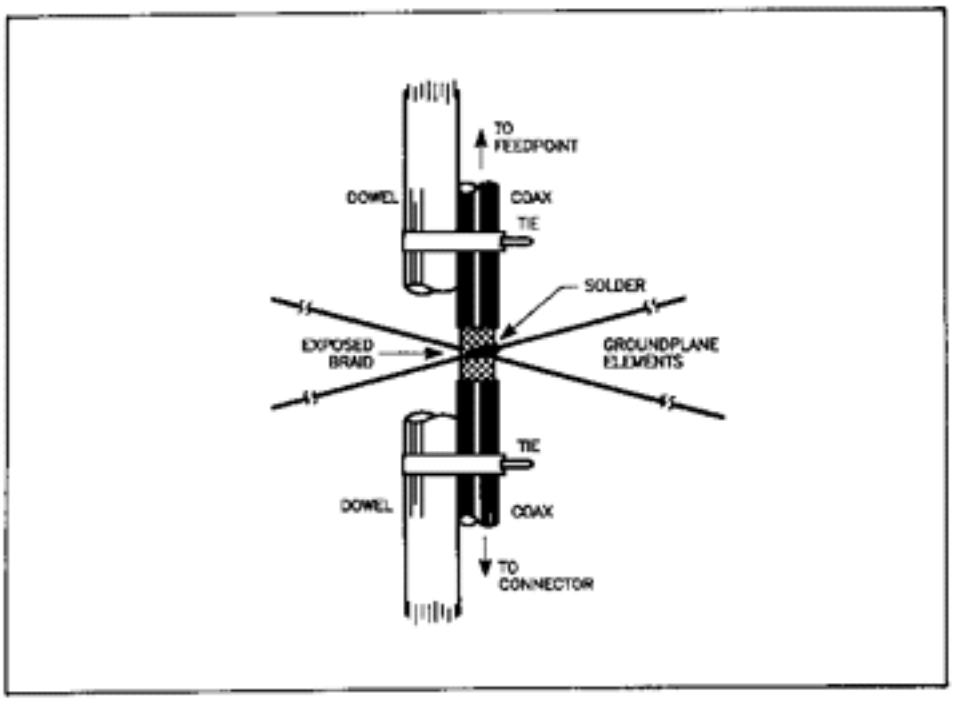

Figure 6. You make the groundplane in this manner. Be sure to thoroughly seal the area of the exposed braid, as described in the text.

though, it's better to pay more and get a cable with the lowest-loss. Belden 9913 is the best of the flexible half-inch cables around and it's stocked by Dick Smith Electronics. You're next best choice would be 'RG213 foam', which is also available from Dick Smith Electronics.

You must mount the collinear well clear of other vertical structures, particularly if they're metallic. The antenna described is readily mounted on a standard TV chimney mount, or even a barge-board mount.

joint to hold it securely to the dowel.

Now, temporarily tie the coax to the dowel near the exposed shield braid. This will secure it while you attach the groundplane elements. Position each groundplane element on the exposed braid and solder them in place at right angles to each other. Take care to solder them properly but not to damage the coax, either (a hot iron with a spade tip is best for this).

13) When the joint is cool, take off the temporary ties and apply a little silicone sealant around the groundplane joint. While the silicone is still plastic, slip the two pieces of heatshrink tubing along to cover the groundplane joint and apply a blast of hot air. Afterwards, cover the joint thoroughly with silicone sealant (you don't want water getting into the coax).

14) Put ties around the dowel and coax, either side of the groundplane, then another tie a little below the groundplane to secure the flying lead, leaving a slack 'kink' in the cable so that

Performance

An eight element array like this has 9dB of gain over a dipole. Your 10 watt rig will sound like an 80 watt rig on a Slim Jim, or like a 100 watt rig on a groundplane – it's cheap gain! A transistor power amp to take your rig's output from 10W up to 80- or 100 watts will cost you $2 per watt, or more. So this collinear costs about one-tenth the price of a power amp. So, how much power will it take? As much as you're legally allowed to run 'up the stick.'

If you live in a valley and hope this antenna will 'get you out', expect the unexpected. It may make things worse because of its low radiation angle. The gain is achieved by compressing the vertical radiation angle. Try it. If you don't get the improvement expected, chop off the top four elements and try

PARTS LIST

Two metres of RG58CU (MIL-C-17F preferred)
500 mm of 9.5 or 12.7 mm diameter heatshrink
100 mm of 6.4 mm heatshrink
in-line coax connector to suit (BNC suggested)
800 mm of 18 gauge tinned copper wire or brazing rod
Two metre length of 12.7 mm or 19 mm diameter dowel
Five or six cable ties
Silicone sealant
Solder
Outdoor wood stain or linseed oil

TOOLS YOU'LL NEED

Sharp penknife or 'hobby' knife
Sharp, pointed sidecutters
Soldering iron, preferably temperature controlled
Small shifting spanner
Pair of needle-nose pliers

groundplane to secure the flying lead, leaving a slack 'kink' in the cable so that any tension is taken by the bottom-most tie.

15) Last of all, put some sort of cap over the top end of the dowel and seal it to prevent it weathering. A damp dowel degrades the collinear's performance, so use a rubber furniture bung of the right size. Or, a short length of heatshrink tubing of the right diameter, tied off and shrunk in place.

That completes the construction. Now to erect it. As individual circumstances vary widely, I'll just give a few hints and tips.

The bottom end of the dowel can be clamped to the top of a mast using hose or muffler clamps that are tightened with a worm-drive mechanism. Use two clamps spaced apart a little to properly support the dowel.

The feedline from your collinear to your rig should be a good quality, low-loss coax. The large diameter 'half-inch' variety is readily available, and affordable. For these frequencies, vertical radiation angle. Try it. If you don't get the improvement expected, chop off the top four elements and try again. It sounds weird, but I know of one constructor who successfully performed this operation, to his surprise, but not mine!

I have made various versions of coaxial collinears over the years, for both temporary, permanent and portable applications. A portable collinear is easily made by tying or taping the coax elements and lead cable to a length of hemp rope. In use, the top end of the rope is tied off to something suitable, like a tree branch or other form of 'skyhook', and the bottom end is either tied down or weighted so that the array is held vertical. When not in use, just roll it up.

I've made four-element coax collinears for 2m, in both 'fixed' and portable versions, an eight-element centre-fed horizontal monster some 15 metres long for six metres, and UHF versions ranging from a four-element job for mobile use to a 16 element phallic symbol nearly four metres tall. ●

Build A 9 dB, 70cm, Collinear Antenna From Coax

By N1HFX

Recently the RASON technical committee was hard at work at the repeater site repairing our 2 meter repeater antenna. One of the members commented to me that I should write an article about collinear arrays so that we could all build our own. While it is not always feasible to home-brew a commercial quality antenna designed to take hurricane force winds, it is very feasible to built a collinear antenna for average use. This article describes a collinear antenna made from very inexpensive RG58/U coaxial cable and encased in PVC pipe.

Before we start building we need to cover some ground about the characteristics of coaxial cable. First remember that there is something called the velocity factor for coaxial cable. For RG58/U coax it is typically .66. This means that when we calculate the length of ½ wavelength in free space we need to adjust its size by multiplying it by the velocity factory. Simply put, RF slows down by the velocity factor when traveling through coaxial cable. All that aside now, lets calculate the ½ wavelength of RG58/U coaxial cable with a frequency of 444 Megahertz:

½ wavelength of coax = 300 / F / 2 * V

Where F = Frequency in Megahertz

V = Velocity factory of Coax

300 / 444 / 2 * .66 = .2229 meters or 223 millimeters

To allow for cutting the ends of our coax, we will need to add 8 millimeters to each ½ wave length for a total of 231 millimeters.

To get started, we will need 8 half wave lengths (231 millimeters) of RG58/U coaxial cable to be cut and connected in the manner shown in Figure 1. First cut back 4 millimeters of the outer jacket, braid and dielectric exposing the center conductor as in Figure 2. Now cut back the outer jacket another 4 millimeters to expose the braid and push the braid back about a millimeter to prevent it from shorting with the center conductor. It is best to lightly tin the braid with solder at this point. Now solder each half wavelength as shown in Figure 1. Attach a few feet of RG58/U to the bottom of the array as in Figure 1 for feeding the antenna.

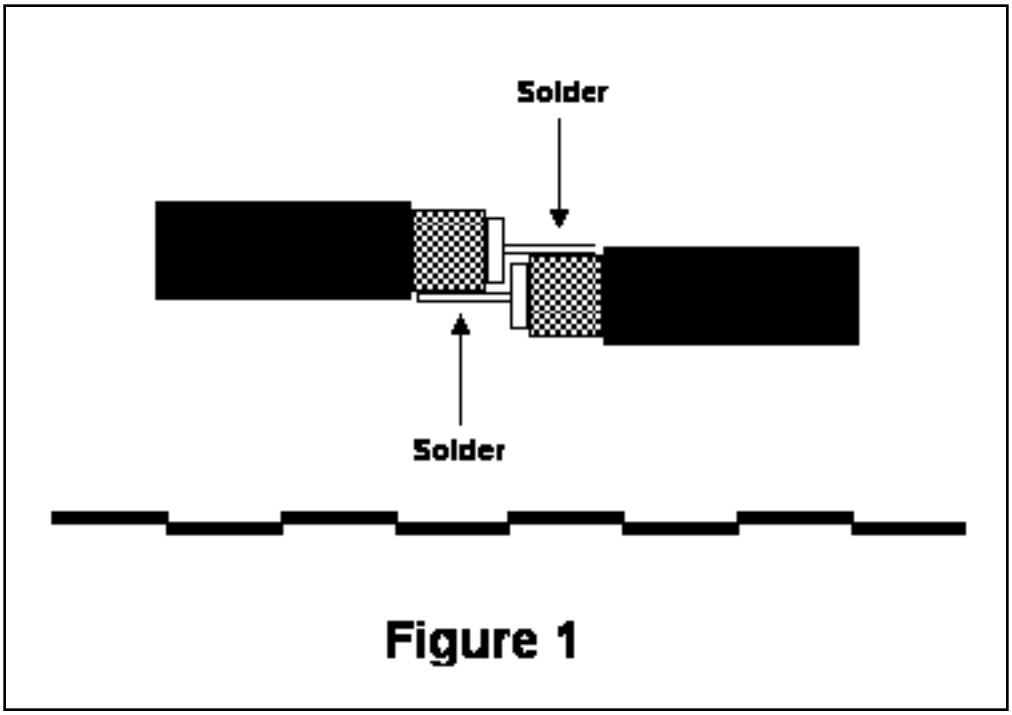

Figure 1

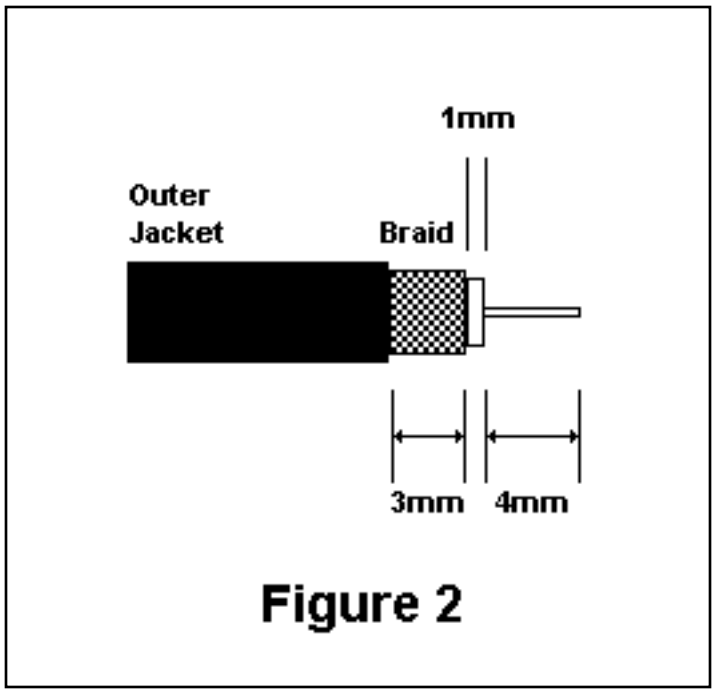

Figure 2

Now its time to add the additional elements to the top and bottom of the collinear array. First add a ¼ wave element to the top of the antenna as shown in Figure 3. Use #16 solid wire or similar and solder it to the center conductor only. The length of the ¼ wave element is calculated as follows:

1/4 wavelength radiator = 300 / F / 4

Where F = Frequency in Megahertz

300 / 444 / 4 = .1689 meters or 169 millimeters

At the bottom of the array we will slide a 5/16 inch aluminum tube over the coax and crimp it to the braid of the antenna feed point only. If copper is used, it is okay to solder. The length of the tube is calculated as follows:

¼ wavelength of tubing = 300 / F / 4 * V

Where F = Frequency in Megahertz

V = Velocity factory of Tubing. (Use .95 for 5/16" tubing)

300 / 444 / 4 * .95 = .1604 meters or 160 millimeters

Because a collinear antenna is hot with RF along the shield of the coax, it is necessary to prevent the RF from coming back through the coax. Slide three FT50-43 or almost any similar sized toroids over the bottom end of the coax as shown in Figure 3. The toroids should be placed about ½ wave length from the bottom of the array. Use the same formula for calculating a half wave length of coax. If you prefer, apply RF to the antenna at this point and slide the toroids up and down until minimum SWR is found. Tape the toroids to the proper point on the coax using electrical tape or similar means.

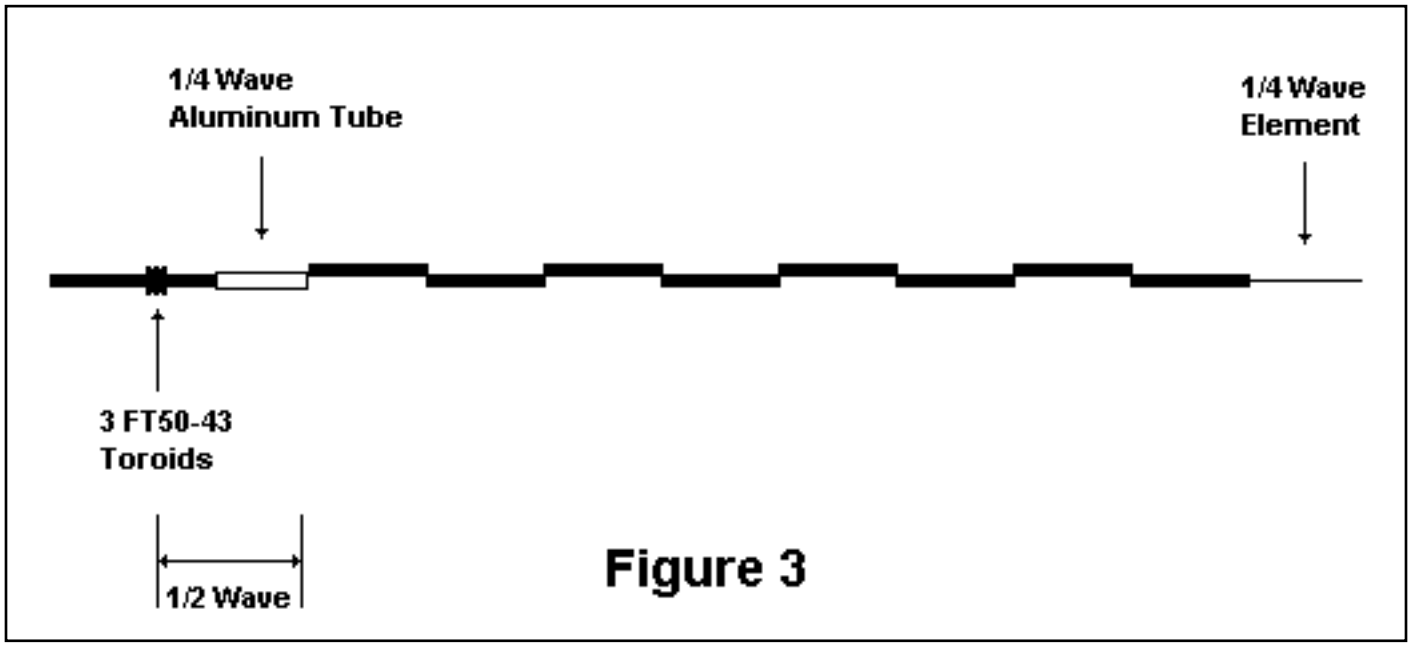

Figure 3

After completing the basic assembly of the collinear antenna, apply a small amount of RF with the antenna on the floor or ground. Relatively low SWR should be observed at this point. The SWR will be much lower once the antenna is mounted in the air. If the SWR is greater than 2 to 1 across the entire band, a connection may separated or a short occurred. It will be necessary to correct the problem before proceeding. After good SWR is obtained, place heat shrink tubing along all connections or wrap tightly with electrical tape.

For final mounting, attach the antenna to a ¼" wooden dowel using tie wraps about every 3 inches. It may not be possible to obtain a wooden dowel for the complete length so attach two dowels together by using a 1 inch sleeve of 5/16" tubing and crimping the tubing at each end. Check SWR again to insure that no connections have separated or shorted. Carefully insert the coax and dowel assembly into several feet of ¾" PVC pipe for final mounting. Because of the tie wraps, it is not necessary to use spacers but may be necessary if larger size piping is used. Drill a hole for the coax at the bottom end cap and place an end cap on the top of the PVC. Do

not cement end caps until the SWR has been doubled checked. Cement end caps and water proof coax opening on the bottom. Use whatever type of coaxial connector is desired on the bottom of the coax end but do not use RG58/U for your complete feed line. Use a low loss coax such as RG8/U for the main feed line to the transceiver. Don't forget to water proof all coax connectors.

If the eight ½ wave coaxial elements result in an antenna too long for your liking (over seven feet), then it is okay to use four ½ wave coaxial elements but the SWR may be slightly higher (Attach four ¼ wave vertical ground radials at the antenna feed point to help lower SWR.). If 9 dB gain is still not enough for you then increase the number of coax elements from eight to sixteen. You will probably need to attach guy lines to the antenna. Although only a 70 CM antenna was described in this article, the formulas can be easily calculated for the 6 meter, 2 meter or 1¼ meter bands. Millimeters were used for many of the measurements but can be converted to inches by dividing millimeters by 25.4 for those who are not familiar with the metric system. After installing one of these antennas, be prepared to hear stations and repeaters that you never heard before.

DE N1HFX

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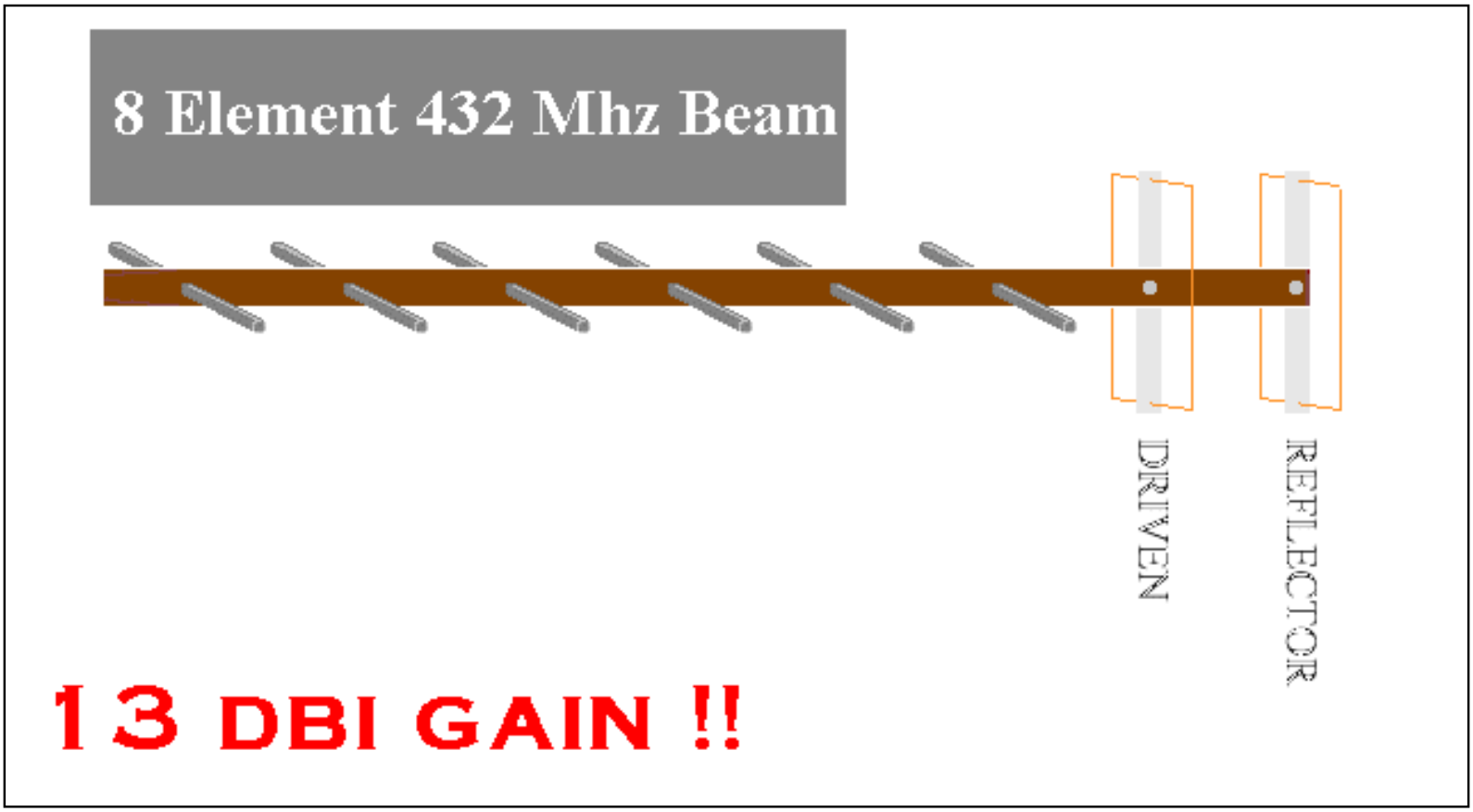
432 8 Element Quagi Antenna



How to build a 432 Mhz Quagi

The boom is made from wood.*{Do not use any type of metal for this.}* Ths boom length is 61 inches and 1/2 inch thick. Mark the boom as to where the elements are to be spaced and drill a 1/8 inch hole in the center of the boom material.After you have done all this apply a few coats of stain.This will preserve the wood from the abuse of the weather.
Use #12 wire to form the quad elements. Cut the wire to the correct lenghts *(see chart below)*. The quad elements are supported at the top and bottom of the element with a Plexiglas strip with a hole centered at both ends. The bottom of the quad being the feedpoint. Apply a little epoxy to where the wire passes through the support holes.This will secure the wire so it will not move around.
Next solder one end of the wire to the center of an type-N conector feed it through the holes in the Plexiglas and bend it into shape. Then finish the element by soldering the loop closed to the ground tab on the N connector.On the reflector just solder to two ends of the wire together.
The directors are mounted through the boom. Epoxy them as well. They can be made

from any 1/8 inch metal rods. Cut them to the EXACT size *(see chart below...I used stainless steel welding rods which are available at a hardware store)*

NOTE: *At UHF frequencies even 1/8 inch difference in length will make the antenna preform differently.*

Element Lengths	
Reflector Loop	**28"**
Driven Loop	**26 5/8"**
Directors	**11-3/4" to 11-7/16" in 1/16" steps**
Element Spacing	
R - DE	**7"**
DE - D1	**5.25"**
D1 - D2	**11"**
D2 - D3	**5.85"**
D3 - D4	**8.73"**
D4 - D5	**8.73"**
D5 - D6	**8.73"**

Click on the banner

432 TECH LIST

MOD---------TECH----------FILES NEEDED

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VHF/UHF MESSAGE BOARD

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HIGH GAIN VERTICAL - COLINEAR MADE FROM COAX (REB) col432.zip

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CONVERTERS

CONVERTER FOR 432 (QST 3-83) fttcv.zip

PREAMPS

PREAMP (VHFER 3-67) c1.zip
FET PREAMP 2MT AND 432 (73 7-67) tfet.zip

TRANSVERTERS

A HIGH PERFORMANCE 432 TRANSVERTER PART 1 (QST 8-91) see below
----------PART 2 (QST 9-91) hfamp.zip

BACK TO HOMEPAGE

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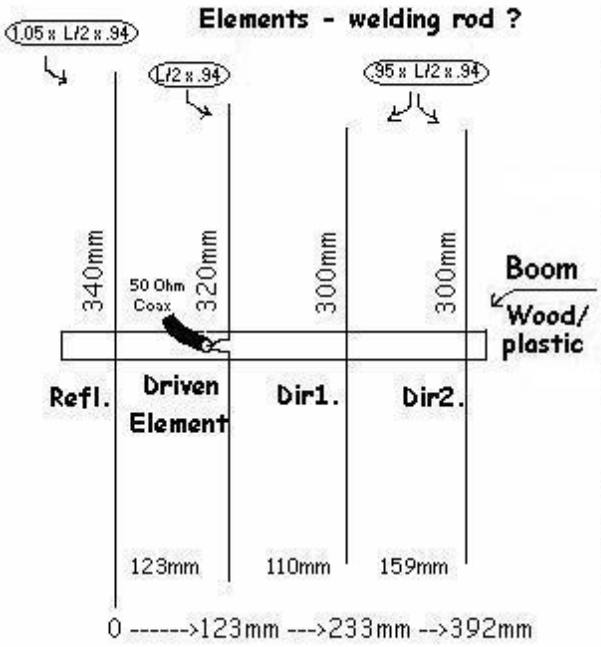
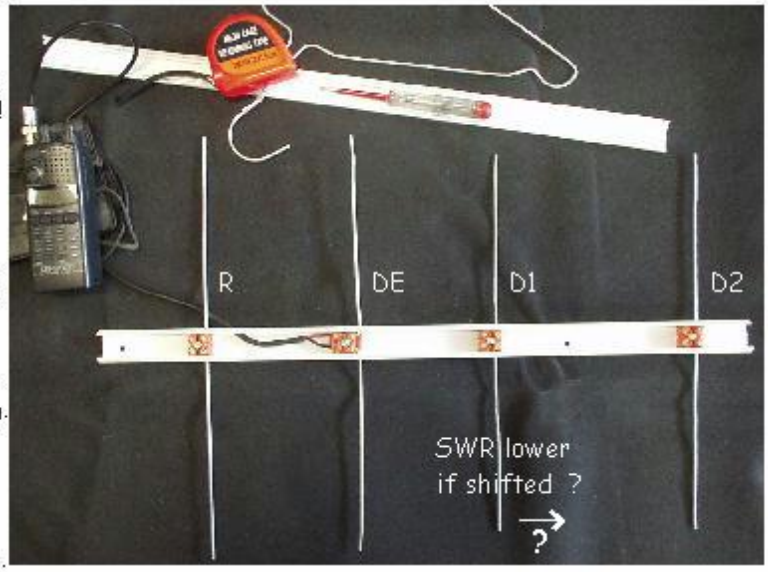


One easy version - uses electrical conduit for boom & galvanised iron coat hanger wire !

Terminal block connector strip makes for great element securing, & neatly connects coax. feed to the 2 parts of the dipole driven els. - each 160mm long.

It all dismantles too- no glue or solder- so suits field use/repairs!



Sourced => www.bk.fr/amatech/electronique/radio/antennes/yagi4/yagi4.htm
Designed under antenna simulator software - "MININEC" => www.emsci.com

433 MHz UHF Yagi antenna

- suitable low power LIPD/ISM "fox hunting"

The 433.92MHz " Low Interference Potential Device" (LIPD) licence free band suits utility devices < 25mW power. A simple 4 element Yagi boosts signals ~ 7dB (= ~ range doubling) + direction finding for nearby signals.

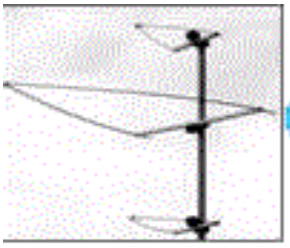
Classic antenna theory, sourced from such Ham texts as the ARRL & RSGB handbooks, assigns the driven element as electrical length =1/2 a wavelength (shown here L) . The actual physical length is somewhat reduced however, by a multiplication factor relating to the diameter of the elements & wavelength of signals in use. At UHF it's typically 0.94. Directors are usually 0.43L, with spacing 0.25L, with improved bandwidth resulting from tapering in the direcxtion of the signal. Gain is not appreciably influenced by rear reflector spacing in the range L/8 - L/4, but useful impedance variation occurs. Hence at 433.92 MHz ...

$L = c/f = 3 \times 10^8 / 433.92 \times 10^6 =$ L = 692mm | L/2 = 346mm | L/4 = 173 mm — This is theoretical electrical length

Actual cut lengths × 0.94 these so L= 650mm | L/2 =325mm | L/4 = 163mm — For physical length driven element

It's apparent thus that the design shown is a compacted version, & better performance may result from spacings & element lengths closer to theoretical. Check many further resources via => **http://ac6v.com/antprojects.htm**

Stan. SWAN => s.t.swan@massey.ac.nz 6/03



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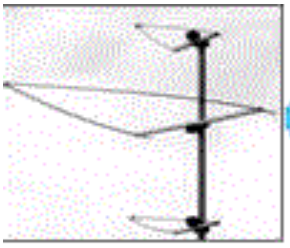

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A 2 - SLUG TUNER PART 1 (QST 1-83) SEE BELOW
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POWER SPLITTER (QST 2-83) psprl.zip
A 1296 SIGNAL SOURCE (QST 3-77) tnsss.zip

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30 WATT LINEAR ON 23CM (DUBUS 2-94) ap.zip
35 WATT 1296 AMP (QEX 6-87) am.zip
1296 AMP - PART 1 AND 2 AND COMMENTS (KF5N) al.zip
1296 AMP (TJ TURNER) aj.zip
LINEAR AMP 1250 (VHF COMM 2-81) ai.zip
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1296 AMP (QST 1-68) 1296.zip
1296 SOLID STATE AMP (QST 12-85) tnss.zip
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1296 FEED HORN ANTENNA (WA5VJB) ac.zip
A QUAGI FOR 1296 (QST 8-81) tnqu.zip
A PARABOLIC DISH FEED (QST 1-83) pdsh.zip
HOME BUILT PARABOLIC REFLECTOR (QST 4-61) hbp.zip
QUAD HELIX ANTENNA (QST 8-63) thex.zip
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3 DIFFERENT PREAMPS (WA5VJB) ae.zip
A LOW NOISE PREAMPLIFIER (QST 6-82) tnlnp.zip
A 1296 PREAMP (QST 1-81) tptns.zip
1296 PREAMP (QST 1-81) tprea.zip

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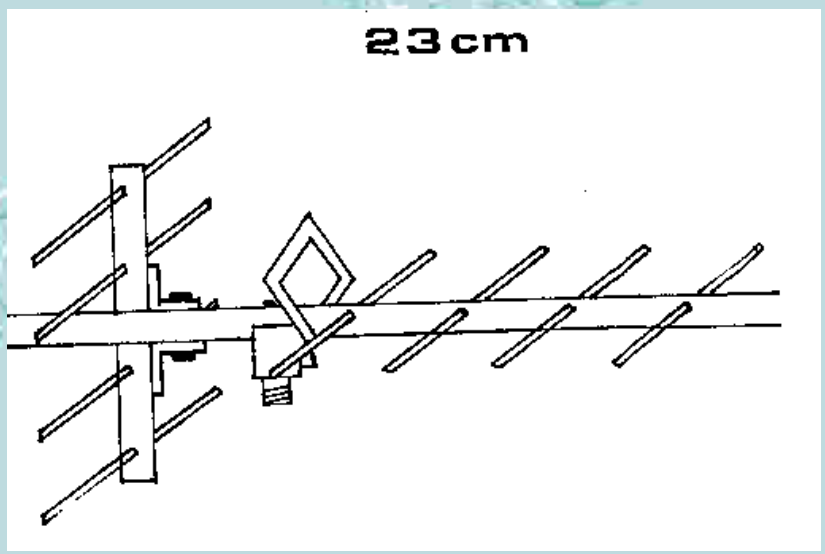
1270Mhz Yagi Antenna

Build your own antenna for the 23cm band (1250Mhz - 1280Mc) using some aluminium and the following simple design.

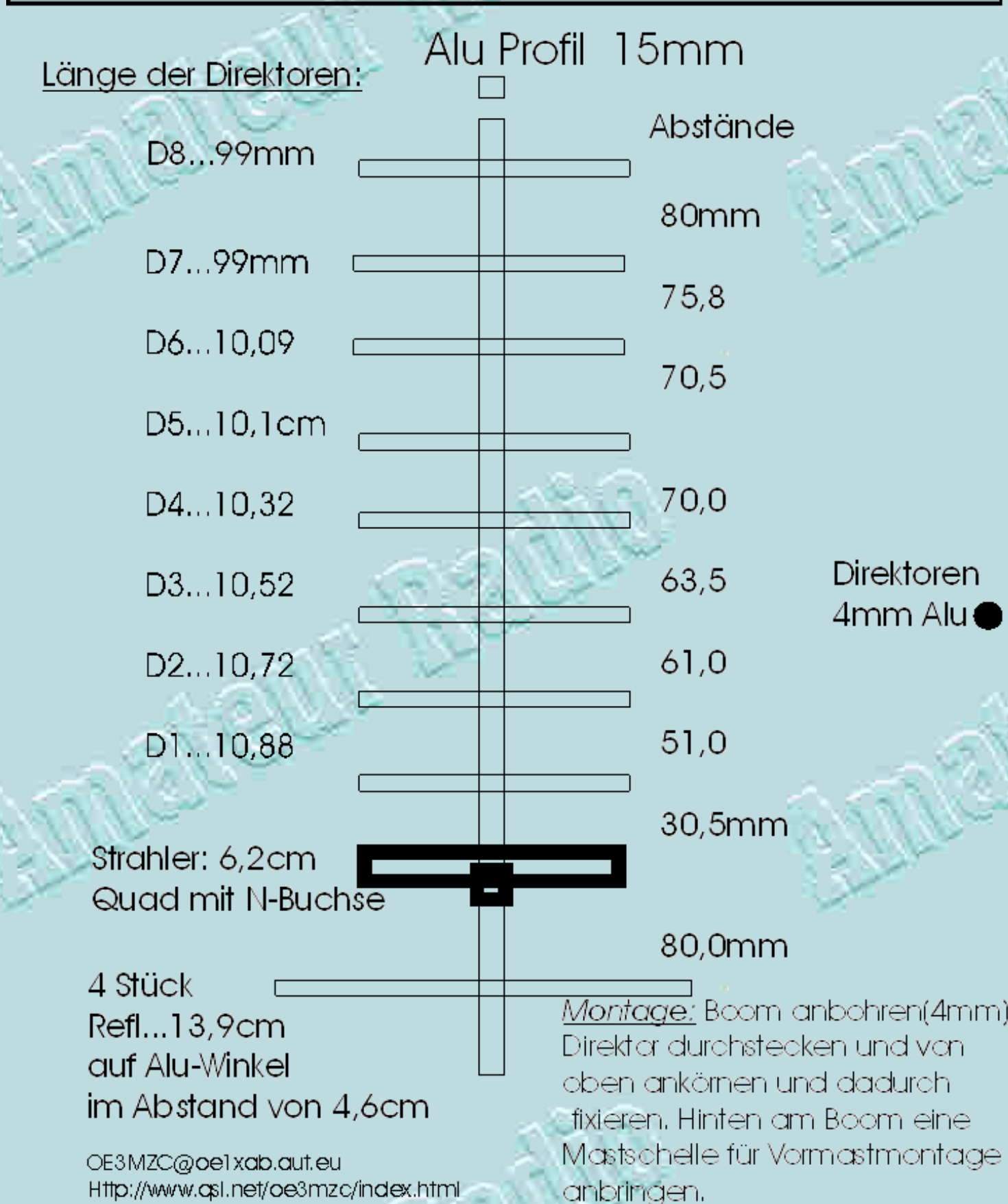


Due to a large number of reflectors front to back ratio is excellent.

The average gain is about +12 dbd.

Broad bandwith makes this antenna the optimal choice for ATV (television)

23 cm Yagi 13 Elemente
ca. 12dbd



OE3MZC@oe1xab.aut.eu
Http://www.qsl.net/oe3mzc/index.html

anbringen.

2,4Ghz - 13cm band quadrible Quad

with reflector 14dbd gain

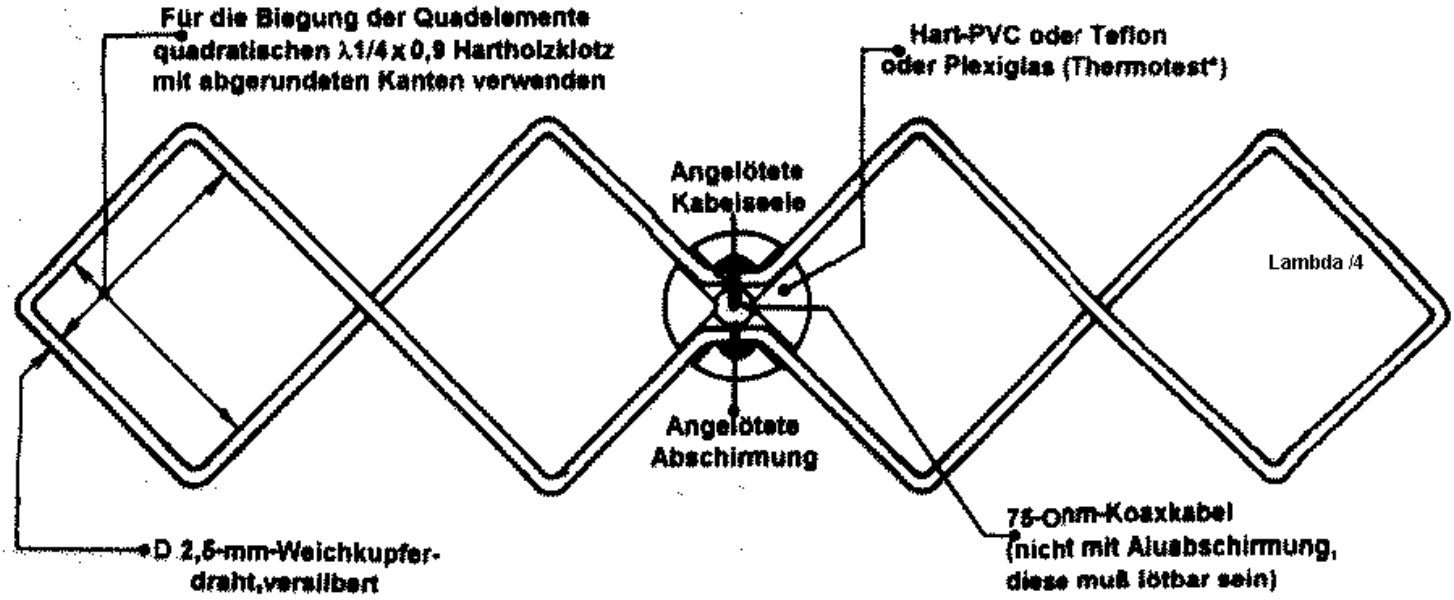
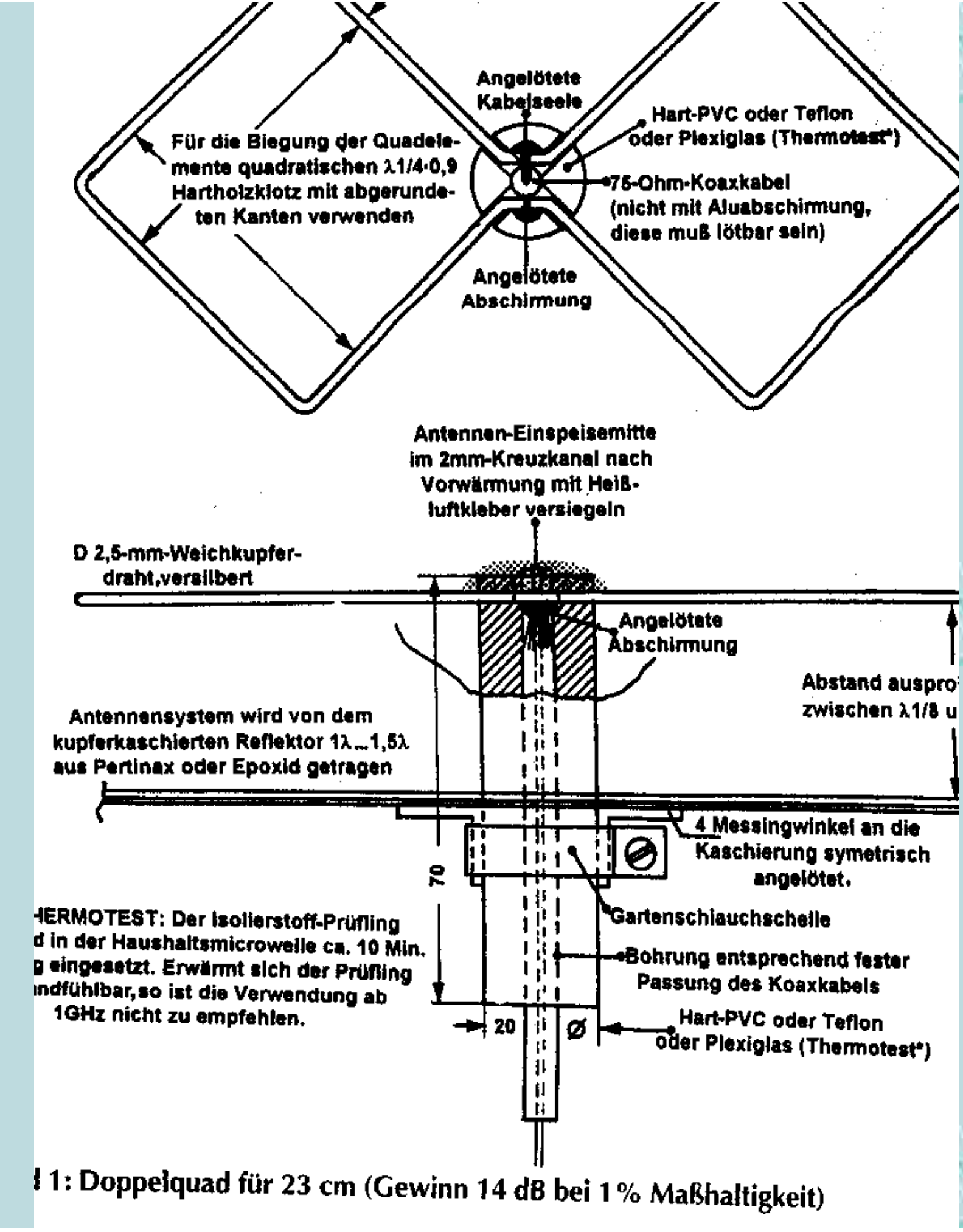


Bild 2: Vierfach-Quad für 13 cm (Gewinn 14 dB bei 1% Maßhaltigkeit, Draufsicht analog Bild 1)

Double Quad antenna

for 23cm band 1,2Ghz

homemade



l 1: Doppelquad für 23 cm (Gewinn 14 dB bei 1% Maßhaltigkeit)

11 Doppelquad für 23 cm (Gewinn 14 dB bei 1% Maßhaltigkeit)

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Greg's Wireless Networking (WiFi) Info Page

- Results of my first Homebrew WiFi Antenna Shootout
- How to Build a Tin Can Waveguide WiFi Antenna
- My Presentation to the Bellingham Linux Users Group (BLUG) Wireless Special Interest Group Meeting
- My Wireless Bookmarks
- Quicktime Video of Pringles Can Antenna from AirShare.org

SardineCanAntenna

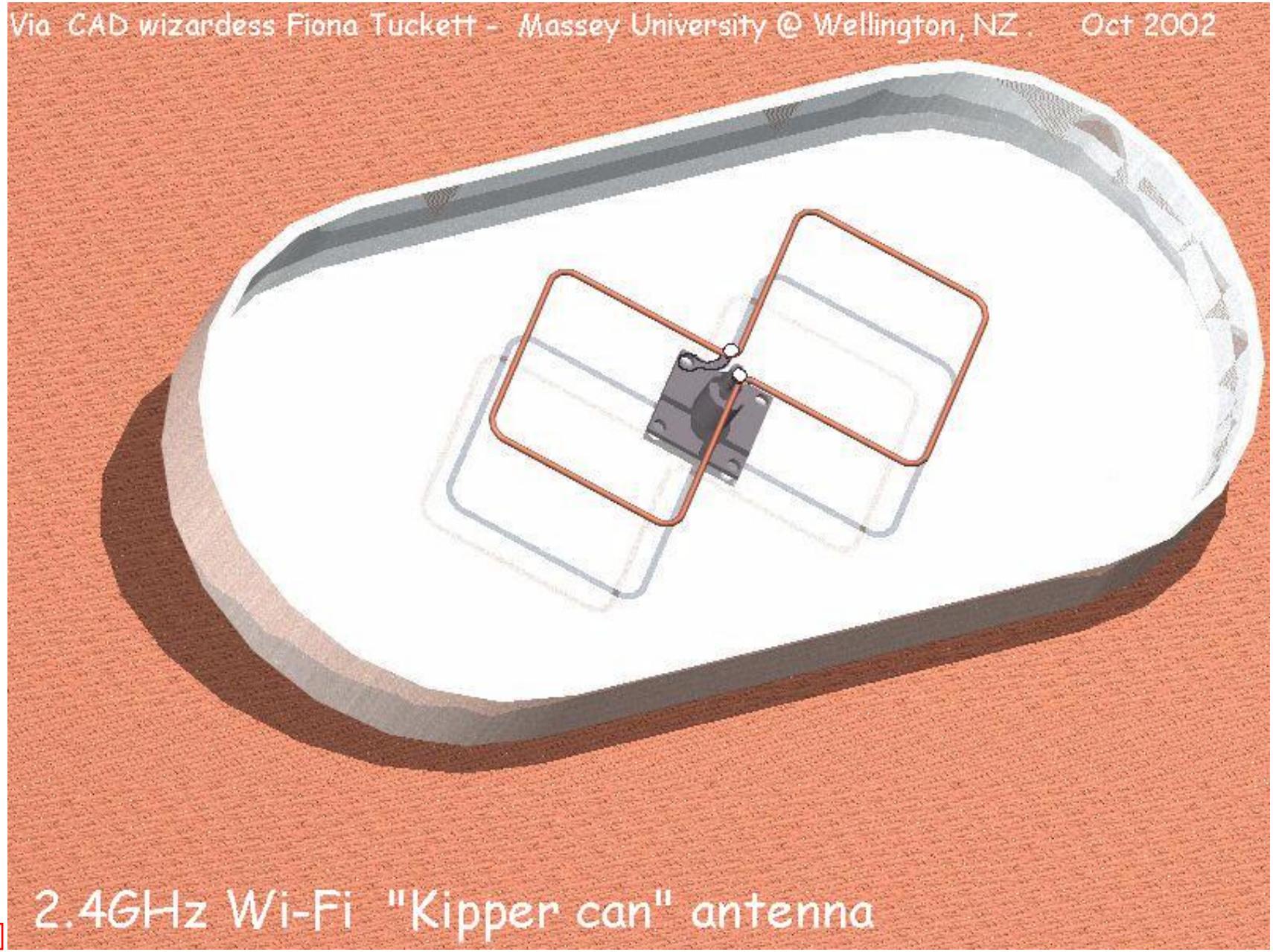
<-- Back to **AntennaHowTo**

Sardine Can Antenna. is a BiQuad or Bi circle - wire length each side (8 x 1/4 waves) 31 mm

Sardine cans have quite standard dimensions around the world, although the oval one shown here is lightweight aluminium & does seem (perhaps by chance ?)to have enhanced directivity compared with more normal steel Canadian "Brunswicks " etc. It's an ex. Latvian RICHTER "Smoked RIGA SPRATS" & perhaps first received publicity in the Nov. 2002 Australian "Silicon Chip" monthly (author Stan. SWAN => s.t.swan@massey.ac.nz), when it was first called a "Kipper Can". (For those readers who don't recall Monty Python, Kippers are smoked herrings http://www.deliaonline.com/ingredients/ingredientsatoz/i_0000000135.asp. See refs & further construction dimensions/views/insights => http://www.manuka.orcon.net.nz <= Even an old CD can act as the reflector, exploiting the metalised layer of course, but the sardine version prooved far superior, no doubt due to the focusing side walls.

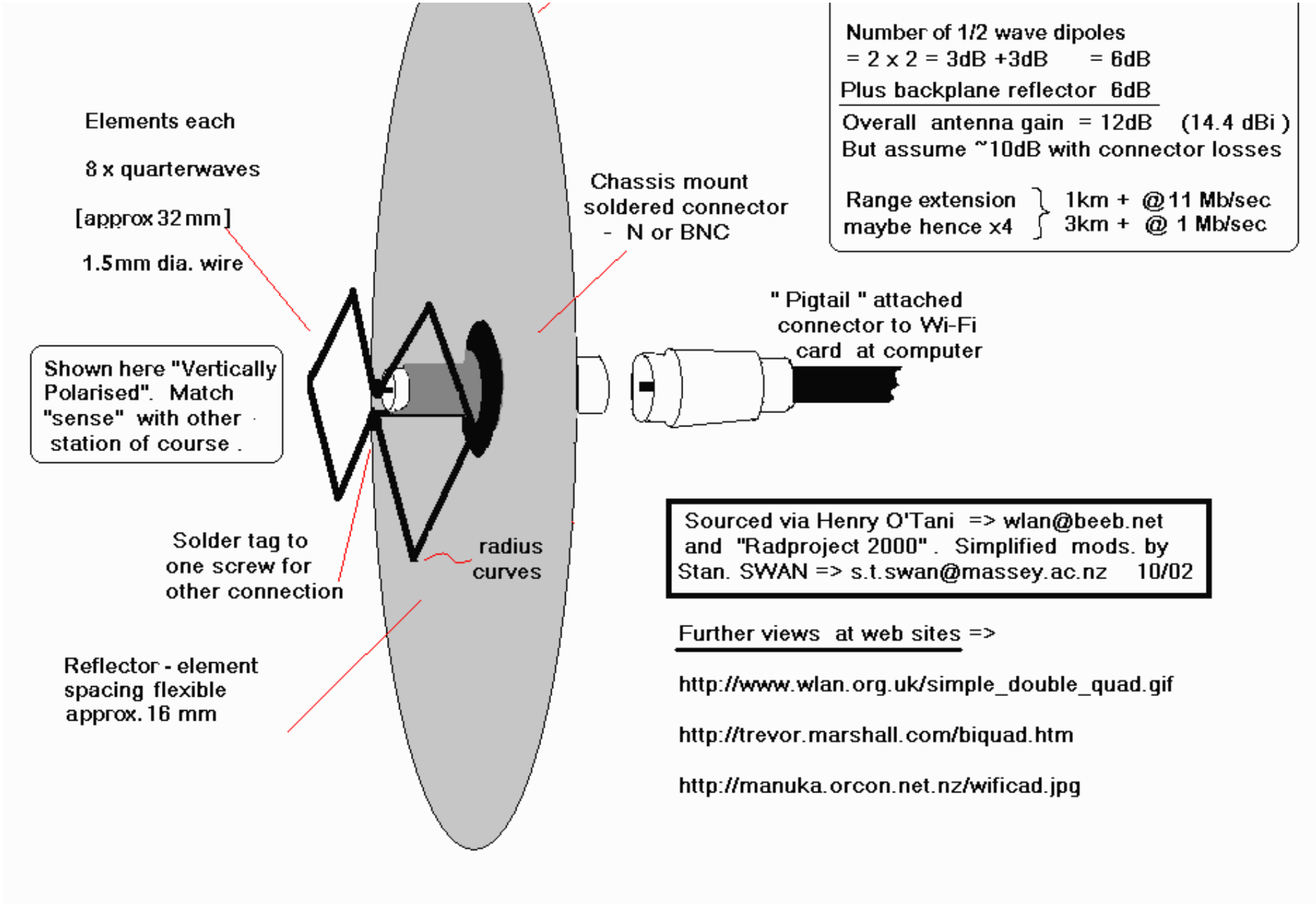
The hot melt glue shown here is not really needed if a rigid N connector is used, although a recent enhancement has been to form the radiating bow tie as tracks on a small PCB secured via nylon spacers. Gain is great -typically 10dB (links to 10km LOS made & typically 2km more normal built up areas !).

Need a quickee antenna ? Got 10 minutes & simple tools ? You won't get a simpler, cheaper or more compact design. The unit can be neatly hidden in a pencil case or plastic bag beside your road warrior notebook so it's not intimidating for field use. It's directional enough to even pin point the signal location (using NetStumbler etc)in a multilevel building. RECOMMENDED !



Double Quad "Bow Tie" antenna for 2.4 GHz Wi-Fi range extension

Gain: (relative to half-wave dipole)



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Construcción de una antena direccional

Loop-Uda-Yagi para 2.4 Ghz.

Por Inco
Fotos : Gonzalo

0001 pelando el ...

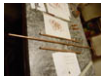
0002 enrollando ...

0003 marcando an...

0004 cortando an...

0005 muchas anil...

0006 varios.jpg

0007 cortando tu...

0008 mesa y tubo...

0009 haciendo ma...

0010 pinza para ...

0011 primera ani...

0012 la segunda.jpg

0013 soldando la...

0014 soldando.jpg

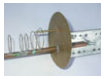
0015 otro soldan...

0015 va tomando ...

0016 reflector.jpg

0101.jpg

0102.jpg


0103.jpg

0104.jpg

0105.jpg

0106.jpg

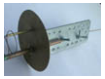
0107.jpg

0108.jpg

0109.jpg

soporte loop.jpg

soporte loop2.jpg


soporte y reflec...

El diseño de esta antena se basa en otros encontrados en internet y sobre todo en los datos obtenidos con el programa de G6KSN loopyagi.exe antes alojado aqui y que sirve para calcular antenas loop-uda-yagi para cualquier frecuencia. Las dimensiones y la forma de construirla se han cambiado levemente para adaptarnos a los materiales que teníamos a nuestro alcance. Es una antena muy direccional y con ganancia bastante alta, 14dbi. La polarización horizontal o vertical depende únicamente de la posición en que fijes la antena.

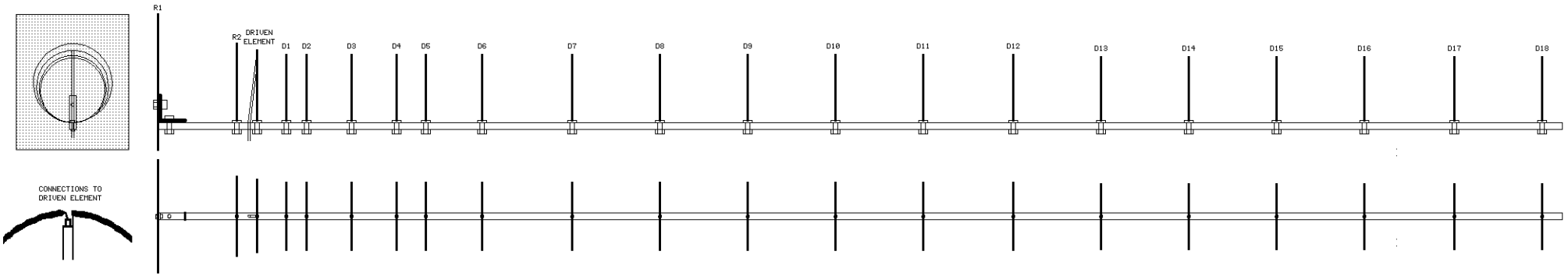


Figura 1: Esquema de la antena original de G6KSN

Los resultados obtenidos con loopyagi.exe para una frecuencia de 2441Mhz son los siguientes :

Elemento	Dimensiones	Distancia desde el reflector1
reflector 1	123mm diametro	0 mm
reflector 2	135mm circunferencia	42 mm

alimentador	123mm circunferencia	55 mm
director 1	114mm circunferencia	70 mm
director 2	114mm circunferencia	81 mm
director 3	114mm circunferencia	105 mm
director 4	114mm circunferencia	129 mm
director 5	114mm circunferencia	146 mm
director 6	114mm circunferencia	177 mm
director 7	114mm circunferencia	225 mm
director 8	114mm circunferencia	273 mm
director 9	114mm circunferencia	321 mm
director 10	114mm circunferencia	369 mm
director 11	114mm circunferencia	417 mm
director 12	114mm circunferencia	465 mm
director 13	110mm circunferencia	513 mm
director 14	110mm circunferencia	561 mm
director 15	110mm circunferencia	609 mm
director 16	110mm circunferencia	657 mm
director 17	110mm circunferencia	705 mm
director 18	110mm circunferencia	753 mm
director 19	110mm circunferencia	801 mm
director 20	110mm circunferencia	849 mm
director 21	106mm circunferencia	897 mm
director 22	106mm circunferencia	945 mm

El radomo (palo al que van soldados el resto de los elementos) es un tubo de cobre, usado en fontanería, de 12mm de diámetro. Los elementos en forma de anilla están hechos a partir de un alambre de cobre de 1.5mm de diámetro.

Dependiendo de la ganancia que queramos conseguir tendremos que hacer la antena mas o menos larga, aquí damos las instrucciones para hacer una de aproximadamente 1 metro de longitud y 22 directores, que da una ganancia aproximada de 14Dbi. Si se quiere hacer una antena de menor ganancia basta con acortarla hasta donde desees, por ejemplo, una antena de 50cm y 11 directores tiene una ganancia de aproximadamente 11Dbi. Como orientación decir que en las pruebas realizadas se obtuvieron ganancias de 7db (la de 22 directores) y 4db (la de 11 directores) por encima de los resultados obtenidos con una antena tipo bote (de 8.7 cm de diámetro y 16.5cm de longitud).

Empezaremos haciendo las anillas, quitamos la funda aislante del cable de cobre, y a continuación enrollamos el alambre sobre un trozo de tubo de cobre de 35mm de diámetro, hasta hacer 12 vueltas completas. Fijamos el alambre con alguna cinta adhesiva y con una cuchilla, dando varias pasadas, hacemos una marca a todas las anillas. Soltamos la cinta, retiramos el alambre del tubo de cobre y vamos cortando cada anilla por las marcas que hemos hecho. Se manipulan las anillas hasta conseguir que formen un círculo completamente cerrado. Estas doce anillas serán los primeros alimentadores numerados del 1 al 12.

El paso anterior lo repetimos enrollado el cable sobre un tubo de 34mm de diámetro, (nosotros utilizamos un tubo de una aspiradora). Sobre este tubo hacemos ahora otras 10 espiras, las marcamos y cortamos. A dos de éstas les cortamos 4 mm, para utilizarlas como directores 21 y 22, las restantes 8 anillas serán los directores del 13 al 20.

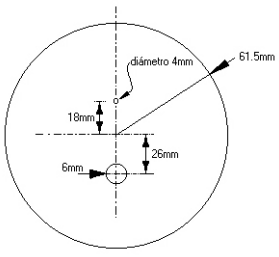
Medimos dos trozos de alambre de cobre, uno de 123mm y otro de 135mm para hacer el Alimentador y el Reflector 2 respectivamente. Les daremos también forma circular enrollándolos sobre un tubo de 40mm de diámetro y rematando la forma a mano.

Cortamos el tubo de cobre que formará el radomo, la longitud depende del número de directores de la antena que nos propongamos hacer. Para la de 22 directores lo cortaremos a 102 cm, es decir, 7.5cm mas largo que la medida que nos indica la tabla (Director 22 945mm).

A continuación sujetamos, con alguna herramienta o cinta adhesiva, una cinta métrica al tubo de cobre y le hacemos las marcas en las que irán soldados los distintos elementos. Comenzamos haciendo una marca a 7cm de uno de los extremos. Esta marca la tomamos como origen o "cero" para le resto de las medidas, o sea, en ella irá soldado el Reflector1. A 42 mm del "cero" haremos la marca para el Reflector2, a 55 mm del "cero" haremos la marca para el Alimentador, a 70 mm del "cero" la marca para el Director1, a 81mm la del Director2 y así hasta llegar al Director22.

Ya sólo nos queda soldar cada elemento en su sitio. Empezaremos por el último director, el 22. Para esto hemos preparado una herramienta o pinza que se puede desplazar por el tubo de cobre y tiene unos brazos que permiten sujetar firmemente una arandela en su posición correcta mientras la soldamos. La soldadura la hacemos con soplete de fontanero, aplicando previamente decapante o flux en las piezas a unir. Tanto los directores, como el Reflector2 se sueldan con la abertura de la anilla en contacto con el tubo de cobre, de modo que al soldar la anilla al tubo queden también unidos los extremos de la anilla.

El Director se suelda de forma que la ranura quede diametralmente opuesta al punto de unión de la arandela al tubo. En los extremos sueltos del Alimentador soldaremos posteriormente el cable coaxial, la malla a uno de los extremos y el vivo al otro. Soldamos a continuación el Reflector 2.



El Reflector1 es un círculo de 123mm de diámetro de chapa de latón de 0.5mm de espesor. Se marca con compás o plantilla y se corta con la tijera para chapa. En este reflector hacemos dos agujeros, uno con centro a 26mm del centro del reflector, y de 12mm de diámetro, en este agujero soldaremos el radomo. Hacemos otro agujero, de 4mm de diámetro y con centro a 18mm del centro del reflector. En este agujero soldaremos un trozo de tubo de latón de 4mm de diámetro y de 60mm de longitud. Por el interior de este tubo se introduce el cable coaxial RG-316, soldamos el cable coaxial al Alimentador y en el otro extremo del cable le colocamos el conector apropiado, dependiendo a que aparato Wifi vayamos a conectar la antena.

Herramientas que necesitas:

- Un cortatubos o un arco de sierra para metales.
- Unas tijeras para cortar chapa.
- Unas tenazillas para cortar los cables, la tijera de chapa puede servir.
- Un soplete, estaño y flux.
- Estaño y un estañador, si es de 100watios mejor que el de 40watios.
- Un tornillo de banco para sujetar las piezas mientras las sueldas.
- Tubos de diferentes diámetros para enrollar las anillas (35mm, 34mm, 40mm)

Donde conseguir los materiales:

- El cable de cobre de 1.5mm de diámetro en cualquier tienda de material eléctrico.
- El tubo de cobre de 12mm de diámetro en cualquier almacén de material de fontanería o inculso en grandes superficies dedicadas a bricolage.
- La chapa de latón la he comprado en Suministros Azán, en el polígono de Argales en Valladolid. Cuesta unos 28 euros una plancha de 100x60cm aprox.
- El tubo de latón de 4mm y la varilla de latón de 4mm de diámetro o similar (para hacer la pinza para sujetar las arandelas) en grandes superficies de bricolage o en tiendas dedicadas a maquetas y modelismo, como Biplano al lado de San Benito, en Valladolid.
- El RG-316 se encuentra en amidata en Madrid, en bobinas de 25 metros.

- El conector N macho para soldar a cable (para rg-58, no se encuentra para cable mas fino) en cualquier tienda de componentes electrónicos, en Valladolid en Oseca, en la carretera de circunvalación. También en Oseca la abrazadera para sujetar la antena a un mástil.
- La chapa perforada para hacer el soporte de la antena se compró en AKI en el Camino Viejo de Simancas, pero puede servir cualquier otra chapa de 2mm de grosor y tamaño 7cmx15cm.

Brought to You by Art Winterbauer (WA5OES)

The Condo Communicator **is a newsletter directed toward helping hams solve the perennial problem of getting on the air from inside apartments, townhouses and anywhere that the traditional big-tower setup is forbidden or impractical. Dormitory residents and anyone not living on 40 acres in the country will find the suggestions in these articles helpful. Enjoy!**

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- The Condo Communicator - Issue 2
- The Condo Communicator - Issue 3
- The Condo Communicator - Issue 4
- The Condo Communicator - Issue 5
- The Condo Communicator - Issue 6
- The Condo Communicator - Issue 7

This has previously been circulated on packet radio and in a modified form on Hap Holly's R.A.I.N. (Radio Amateur Information Network). Enjoy!--WA5OES

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"But I Never Agreed to That!"

Covenants, Conditions and Restrictions in the Amateur Service

By Chris Imlay, W3KD
ARRL General Counsel

From December 1995 *QST*

PRB-1 · Regulatory Information Branch

- How Do Covenants Differ From Zoning Ordinances?
- So Who Enforces These Covenants?
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Jim Rich, WD6CJB, woke up one morning to find that, during the night, someone had accidentally crashed a car through his fence. That wasn't much of a problem all by itself--the insurance company would cover the fence repair--the unfortunate part was that Jim's multiband HF dipole antenna that was mounted on the inside of the fence, was destroyed. The dipole had not been effective for what Jim Rich wanted to do in Amateur Radio--emergency and public service work--so he figured that it was time to put up something better anyway. That decision began the nightmare that Jim didn't have the night before.

Jim And CC&Rs

Jim went to the zoning authorities for the City of Foster City, a group of subdivisions in a planned community south of San Francisco, and asked for a building permit for a 35-foot crank-up antenna support and a small triband Yagi antenna. He was promptly denied such; the mayor and council members went out of their way to keep this "scourge" out of the neighborhood. They were not successful, because the zoning ordinance in Foster City was plainly preempted by

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the FCC's Amateur Radio Preemption policy.[1] The thing about Foster City, though, was that when it was first formed, there was not any zoning ordinance (or local government) in the town. Instead, all of the subdivisions had identical declarations of covenants--often called deed restrictions, or "CC&Rs" (covenants, conditions and restrictions)--that governed land use in each subdivision. When the town adopted zoning ordinances to deal with land use details, the homeowner's associations in the Foster City subdivisions gradually faded away. They were no longer necessary.

The covenants, although dormant, were still in place, and they bound each successive purchaser of land in Foster City. The covenants provided, as do most, that any home-owner could enforce them against any other homeowner. And in Foster City, the CC&Rs prohibited all antennas of every type that are outside the house, above the roof, and not under the eaves. The restriction is absurd, as a practical matter--there are television antennas on numerous houses. When the City of Foster City found out that it couldn't prohibit Jim Rich's modest antenna system, Jim got his building permit. However, the municipal government did all it could to encourage the formation of a group of neighbors (most of whom can't even see Jim's antenna) to enforce the long-dormant and unenforced covenants against Jim. Thus began about four years of litigation. Jim's case ate up thousands of dollars of Jim's money, caused untold emotional impact on Jim and his wife, and resulted in many hundreds of hours of donated time by two very dedicated and talented ham lawyers in northern California [2] in the efforts to protect Jim's ability to install and keep a modest 35-foot crank-up antenna.

How Do Covenants Differ From Zoning Ordinances?

Many hams are not aware of the differences between zoning ordinances and covenants. The two are radically different in terms of their legal status and how they are likely to affect an amateur. It is useful to understand covenants not only from the viewpoint of the ham who appears to be at the mercy of his or her neighbors, but as well as from the point of view of the real estate developer, who has created the problem for the radio amateur in the first place.

Suppose you are a real estate developer, who looks forward to a financially successful subdivision development project. Financing a subdivision and the construction of houses requires that you borrow money from a bank. You then acquire the land for the subdivision, file a subdivision plat with the county or city, and begin construction of houses. Your interest is in getting your houses built and sold quickly, so you can pay off the bank, pocket your profits, and move on to the next project. The process of selling off the houses in your subdivision, however, may not go as fast as you would like, and perhaps you have to keep a few houses for some time. Since the last few houses sold represent, for the developer, the profit in the subdivision project (the bank has to be paid off first), what you don't want is for one of the earlier purchasers of a house in the subdivision to paint pink camels on the front of his house before you can sell off all remaining houses. Some potential home buyers might find the camel paintings unattractive. That would make it difficult for you to sell the last few houses. But what can you do to prevent the pink camel painting? You already sold the painter his house. It isn't yours anymore, and the zoning ordinances don't address pink camel painting.

The answer comes from the old English common law, from where have derived our law and procedures for property ownership and land transactions. Covenants have been placed on land for many years, and have evolved as a means of private land use restriction. Some covenants, in addition to regulating the use of the land itself, seek to regulate even the behavior of those who own the land.[3] Covenants are created when the developer files, with the land records, and usually together with his or her subdivision plat at the time land is subdivided for development, a declaration of covenants. These restrictions on the use of land are applicable to all parcels of land in the subdivision. So, just by buying the land, a buyer becomes subject to the restrictions applicable to the property bought.

Simply stated, covenants are promises that the buyer of land makes as a condition of purchasing the land in the first place, from either a developer or a previous owner. Because the covenants are filed with the land records in the county or city where the land is located, and because they are referred to in your deed, you aren't even required to be fully aware of the specifics of the covenant restrictions when you go to closing on the land.[4] Since they show up in the "chain of title" to your land, the buyer of the land has what is called "constructive notice" of the covenants, and therefore he or she is charged with the knowledge of them. In other words, you are expected to know what is in the covenants, even though you may not have actual knowledge of the restrictions in them. In many states, this is enough all by itself to require that the buyer comply with all of their provisions. The situation is another expression of the concept that "ignorance of the law is no excuse."

So Who Enforces These Covenants?

Initially, the developer retains the rights to enforce the covenants, since they are to his or her benefit in the first place. Remember that the developer wants the homes in the subdivision to pretty much stay the same, in terms of appearance, until he or she can sell off all the houses and take his or her profit and go on to the next development project. After that, the normal provision is for the enforcement authority to be delegated to a homeowner's association or an architectural control board formed by the declaration of covenants. Sometimes these associations exist for a while and then disappear. If there is no association, does the enforcement authority disappear as well? Sadly, no. In most states, individual homeowners, who are supposedly benefited by the covenants, also have the right to enforce the covenants as well.

Antenna Covenants

Covenants regulating antennas come in various forms. According to studies commenced by creative communications lawyer and professor Wayne Overbeck, N6NB, antenna covenants take three typical forms: (1) covenants that prohibit all outdoor antennas, (2) covenants that prohibit all radio transmitting devices on the premises, and (3) those that require, prior to installing any "structure" outside the house, the approval of the homeowner's association.

Suppose you buy a house in a subdivision subject to CC&Rs and then become a ham and want to put up an antenna. If you can't get the approval of the association to do so, what do you do? Simply putting up an antenna in violation of the covenants is not necessarily a good plan, because many homeowner's associations, funded by regular, mandatory dues from homeowners in the subdivision (as provided in the declaration of covenants) are charged with the enforcement of the covenants, and are financially able to bring a lawsuit against you to do just that. Indeed, if they don't bring an enforcement suit against you, other neighbors might sue the association for failing to do its job. The worst part about it is that, if you are sued by the association and lose the case, or if you sue the association and lose, attorney's fees incurred by the association (and occasionally, fees incurred by individual home-owners) are, according to many covenants, assessed against you. If you don't pay the fees, the debt continues to haunt you in the form of a lien on your property.

There are common-law defenses against covenants, but they are not used very much, and their effectiveness has been reduced over the years. However, if there are multiple violations of an antenna covenant in a neighborhood (such as the unchallenged existence of other amateur antennas, CB antennas, satellite dishes, or television antennas), then that is good evidence that the covenants prohibiting antennas are abandoned or waived. If you are told by the association that an antenna installation will be acceptable, that approval, if you can prove that it was given, makes enforcement of the antenna covenant difficult for the association later. And if you have an antenna up for a number of years in violation of the covenant, and if it is visible and obvious to others during that time, it is unlikely that the covenant could be enforced at a later date. But beware of these defenses, as they are not absolutely available. Jim Rich noted in

his case that there were antennas in the neighborhood, but the judge decided that they were minimal by comparison to his 35-foot crank-up tower and standard triband Yagi, and did not constitute a waiver or abandonment of the covenants.

How About PRB-1?

Nor does the FCC's PRB-1 preemption order apply to covenants. It specifically disclaims any application to covenants, which FCC describes as "private contractual agreements." It said that "Such agreements are voluntarily entered into by the buyer or tenant when the agreement is executed and do not usually concern this Commission."[5]

Why would the FCC conclude that PRB-1 did not apply to covenants, while it did apply to zoning ordinances? Surely, covenants that preclude antennas have the exact same effect on the strong Federal interest in promoting Amateur Radio communication as do zoning ordinances. Remember, though, that zoning ordinances are actions of local government, imposed to protect the health, safety and general welfare of the public. State and local laws may be preempted by the Federal law, where the state laws conflict with the Federal law, according to the Constitution of the United States. Covenants, on the other hand, are not governmental actions. They are, in theory, private "agreements" between a buyer and a seller of land. Therefore, they are insulated against intrusion by the Federal government.

There are exceptions to this, however. Racial covenants were declared invalid in 1948 in a famous Supreme Court case called Shelley v. Kraemer.[6] There, the Supreme Court said that where the state courts were called on to enforce a covenant, those covenants that violate a fundamental right should be held invalid. If the FCC had said nothing about covenants in the preemption policy of PRB-1, it might be argued that the "strong Federal interest" in amateur communication is sufficient to support preemption of covenants. However, the Commission's disclaimer of an interest in covenant preemption has made that argument difficult.

The FCC explained its rationale for avoiding the preemption of covenants in PRB-1 by saying that "Purchasers or lessees are free to choose where they wish to reside where such restrictions on amateur antennas are in effect or settle elsewhere." Is that so? Tell that to a person attempting to purchase a home in any of hundreds of other cities in the United States. Studies by Overbeck, N6NB, and others showed that virtually all new housing developments in metropolitan Los Angeles, for example (most less than 20 years old), were subject to antenna covenants. The assumption of the FCC (and of the courts that routinely enforce covenants now) is that it is possible to purchase a home without covenants, so the purchase of a home with covenants is a voluntary act. The assumption fails, however, where the covenants cannot be avoided, subjecting the amateur to having to purchase a house with covenants, or else endure a two-hour commute to work each morning in order to pursue Amateur Radio. There simply is no choice in most metropolitan areas these days, and the situation is getting worse all the time.

What's in the Future?

FCC is now reevaluating its land use preemption policies. It is looking at satellite dish preemption again, and soon will consider preemption of local restrictions on cellular and PCS antennas. FCC has promised to revisit its amateur preemption policies at the same time. It would be reasonable in that context for the Commission to state that it has no less interest in the effective performance of an amateur station merely because it is regulated by covenant restrictions rather than zoning, and at least allow the amateur to make his or her own argument, in the courts, that certain covenants may be preempted.

Until then, the best solution to the covenant problem is to avoid purchasing a house with deed restrictions, or make your offer on the house contingent on your ability to install a reasonable antenna. Put the burden of getting that

approval on the seller. If you are already in a subdivision with deed restrictions, get on the board of the home-owner's association and use some politics. Meanwhile, the League will continue to work for some regulatory reform to deal with antenna covenants and other deed restrictions. If you have a covenant problem, let the Regulatory Information Branch at Headquarters know about it.

And What Happened to Jim Rich?

Just before press time, and after over seven years' struggle, Jim's antenna case is finally over. A settlement was reached on October 10, whereby Jim will pay the plaintiffs $6,000 of the $13,000 in fees awarded by the trial judge, both sides will dismiss their appeals, the plaintiffs will release the lien they had on Jim's home, and plaintiffs will not complain about the wire antenna Jim has strung along his fence at a height of five feet.

The trial judge determined that the FCC had the ability to deal with the CC&R issue, but that they had ducked the issue with a footnote in PRB-1. The judge concluded that is was not appropriate for the court to overturn well-settled land law use to address something the FCC itself was unwilling to address.

Rusty Epps, W6OAT, who supplied this late-breaking information, commented that "No one is happy with the settlement terms, so that probably means that it's a pretty good compromise solution!"

Notes

[1] See, PRB-1, 101 FCC 2d 952 (1985), codified at 47 C.F.R. §97.15(e). See also, Chris Imlay, N3AKD, "PRB-1, Seven Years Down the Road," *QST*, Oct 1992. pp 35-37.

[2] The trial law skills of Kip Edwards, W6SZN, and Rusty Epps, W6OAT, and their amazing tenacity and generosity, are stories in themselves.

[3] A good example of this is the prohibition of parking recreational vehicles or boats, or even cars with antennas, in the driveways of houses in subdivisions with CC&Rs.

[4] This is changing by statute in many states; now, it is often required that a buyer of land be given a copy of the CC&Rs prior to closing, at least when buying from a developer.

[5] *Amateur Radio Preemption*, *supra*, footnote 6.

[6] 334 U.S. 1 (1948). For a good analysis of the arguments in favor of Federal Preemption of CC&Rs governing antennas, see Sid Leach, Esq, K5XI, "Federal Preemption of Deed Restrictions," *The DX Magazine*, May/Jun 1993, pp 36-40.

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Welcome! I have been searching the net for months now, looking for legal resources that hams can use for that "tower project". I have all the documents contained in the ARRL "legal packet", and was hoping to find them posted somewhere. With the exception of PRB-1, I have not. As a service to fellow amateurs, I have decided to post them here. Some of my own ramblings and experiences can also be found here. I welcome your comments, suggestions, questions, and answers. If relevant, they will also be posted here on a regular basis.

-Good luck with that tower project! 73, de Gregg, K3QK

-My own experiences with towers and local government officials

-Chronological List of Landmark Amateur Radio Case Cites

-Memorandum Opinion and Order in PRB-1 (from ARRL Web Site)

-PRB-1, The FCC's Limited Preemption of Local Ordinances (from ARRL Web Site)

-John Thernes vs. City of Lakeside Park

-John Thernes vs. City of Lakeside Park-Consent Decree, Order, and Final Judgement

-Andrew B. Bodony vs. Village of Sands Point

-Sylvia Pentel vs. City of Mendota Heights

-Appeal of Lord (Supreme Court of Pennsylvania, 1951)

-Stay tuned! More to come, as soon as I can type it in!

Good Tower/Antenna Sites on the Web

(I am constantly on the lookout for more. Know of one? Email the URL to me!)

-Frank, VE3FLF's, web site Solid info and great photos!

-Don Stoner's Restrictive Antenna Covenants Home Page

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-this page first posted Sunday, April 5, 1998 (more to come)

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CC&R Survival Series

Bill Fisher (W4AN) on January 8, 2001
View comments about this article!

I would like to propose the start of a series of articles from people who are still able to be active hams while living within restrictive covenants. I happen to be one of those people, and I am on the air daily on several bands with good signal reports and long QSOs. I am currently writing an article on an antenna that I am building and hope to have it posted in a week or two.

If you are successfully operating the radio under restrictive covenants and have advice or interesting success stories, please take a few minutes to pass them on the rest of us. If you know someone else who is surviving CC&R's, please drop me a line with their email address and I will contact them about writing an article for us.

73

Bill Fisher, W4AN w4an@contesting.com

Member Comments: Add A Comment

CC&R Survival Series Reply
by AD6LR on January 8, 2001 Mail this to a friend!

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Interesting topic. One item to include with CCR solutions would be how RF exposure issues are resolved; for example, it would be very difficult to have an antenna radiating 1.5 KW (say in the 10 meter band)located in the atic with an operator located close by and still meet the FCC RF exposure requirements.

AD6LR

CC&R Survival Series Reply
Anonymous post on January 8, 2001 Mail this to a friend!

I live in a college dorm apt complex, with blanket antenna restrictions. I have (so far) gotten away with a run of magnet wire from my 2nd floor window to some trees about 50 feet away for SWL work, and I put a 2M groundplane on top of a homebrew T shaped PVC assembly with some pipe hangers and hung it off the gutter outside my window. The restrictions in our housing contract are written so broadly, that they could technically ban cordless phones, cell phones, stereos, clock radios, TVs, walkmans, etc.. (of course this was not the intent I'm sure).

RE: CC&R Survival Series Reply
by W4AN on January 8, 2001 Mail this to a friend!

BTW, I was hoping we could have some articles from you survivors and not just comments posted to this article.

When I was in college, I snuck a piece of coax out the dorm room window up on to the roof. Luckily I was on the top floor. I put up a 40 meter dipole antenna off of the TV antenna mast. Antenna was probably a good 100' off the ground and worked VERY well.

Of course, my buddies with the stereo gear didn't appreciate it. But they were none the wiser of my being a ham. :)

73

RE: CC&R Survival Series Reply
by K1IR on January 9, 2001 Mail this to a friend!

A distinction that would be interesting to understand in each situation is whether the operation is OVERT or COVERT. Is your setup known to management and approved, known to management but not approved, or unknown to management? By separating situations into these categories, a variety of interesting strategies can be devised.

I only operated 2fm from my dorm room in college - and my mag-mount 1/4-wave vertical was confiscated more than once. It was an engineering school. We had a nuclear power lab on campus. We had biotech engineers cloning each other. Didn't management have anything better to worry about than my 2m whip? Good thing we had a great ham station and I was eventually the Station Manager.

73,

Jim

CC&R Survival Series Reply
by K3AN on January 9, 2001 Mail this to a friend!

I chose, and my wife accepted, a home with trees and undeveloped space close behind the house. If you have such "natural antenna supports" on your property, it makes getting on the air in stealth mode much easier.

The end-fed antenna is 130' of #26 black insulated wire in the trees, worked against a single ground rod. The system is matched using the great SGC-230 remote antenna tuner. By using a wire that's a multiple of a half wave at 80, 40, 20, 15 and 10, the feedpoint impedance is relatively high so ground losses from using just the single ground rod are relatively low.

The SGC is fastened to the "back side" of a tree trunk just above ground level, and disguised by setting some broken limbs alongside it to break up its outline. The coax and 12 volt power cables are buried in a shallow trench to the point where they are routed up a deck support post and into the basement

through the sill plate. I laid out and buried the cables one evening when I knew the nearest neighbor was away.

The antenna wire is very difficult to see against the trees, even when you're looking for it. At a previous, rented QTH, I ran this kind of wire from the second story of my townhouse across the parking lot and into a tree. None of the neighbors ever noticed it until one neighbor had a visitor who was a ham. The ham pointed it out, but since my QRP operation hadn't been causing any QRM, the neighbor didn't raise any kind of fuss.

A slingshot with a 1 oz lead sinker can fire 10 lb test line a remarkable distance up and out. I use a fluorescent green line to make finding the far end easier. Cut off the sinker, tie the antenna wire to the line, reel it in from your launch point, and you're well on your way to stealth hamming.

CC&R Survival Series Reply
by WA6KGI on January 9, 2001 Mail this to a friend!

Hello Bill,
Sure I will comment. I am currently buying a house in Texas. It has a deed restriction and I think I can beat it. At least I am going to try because the Homeowner's association does not charge any dues. The restriction is poorly worded too. I am going to first speak with attourney buddies to see what I can do about having it removed from the deed (when I get that as I have not yet seen it.) The land size is nice, about an acre and well treed. I think I can easily hide wires in the trees. A tower is the big question.
I will keep you informed.
73 de WA6KGI
Jack

CC&R Survival Series Reply
by KA4JNB on January 10, 2001 Mail this to a friend!

We had our house built in a very restrictive neighborhood. I particularly hate the restriction against growing a vegetable garden. However, the covenant against antennas is vague, reading "no unsightly structures" which eliminates towers of course.

I planned ahead. I selected a lot with dense woods extending over 300' behind the house. Then I borrowed a plane-table, a fairly simple surveying device, and mapped all the significant trees growing there. With this done, I selected trees to support my dipoles, out of sight, and was lucky enough to have a large maple tree, 150 feet back in the forest, with the main truck forming an arch at about forty-five degrees. Throwing a line over this "limb" and clearing all growth underneath, I pulled a full wave two-element Lighting brand quad wired for five bands up off the ground eight feet. I then poured a 24X24 inch pad with a pipe flange secured to its surface with spikes, screwed a 3 foot galvanized pipe into it, mounted a CDE rotor on the pipe, and used a telescoping Radio Shack mast to connect to the boom. The limb holds the weight and the rotor works great. I painted the poly arms green and brown, and you really have to make an effort to see the setup. (we are in the lowest elevation on the subdivision, and I still compete successfully in the pileups, 5BDXCC, 5BWAS, etc. Not as easy as with a tower at a higher elevation of course, but I still enjoy good DXing)

CC&R Survival Series Reply

by WA4CNG on January 10, 2001 Mail this to a friend!

I am still surviving after making Y2K-DXCC in October last year with my Attic Antennas (Article posted here). You and I live less than 15 miles apart in Alpharetta Ga. I have made some changes up there, got rid of the Square Loops, added a 6 meter dipole. Hamshack and shop is now in the Terrace Level, added an Amertron Al-811H amplifier, feeding the antennas with 1/2 inch hardline in the attic via the conduit I had installed by the builder. I am still working on getting 75 meters to work. I am having mixed results, mostly not, from the DX-SA Antenna I

installed for 75/20 meters. VHF/UHF is great with the ground elevation of 1140FT, plus 35 feet into the attic. That is nearly 1200 Feet ASL in Atlanta Ga, where the average terrain is 1000 feet, like having a 200 ft tower in the backyard.

CC&R Survival Series Reply

Anonymous post on January 10, 2001 Mail this to a friend!

My subdivision bans antennas. I am taking a two prong tact: First, I volunteered for the architectural review board and wound up as chairman. This gives me a lot of "inside influence" especially when convincing others that new PRB-1 actually applies to ham antennas in CC&Rs. Second, I applied a loophole that allows "satellite antennas" (intended for tv of-course) and called my 6 meter rotatable dipole a "satellite antenna" who knows, it might pick up a satellite signal! Since we have some trees, I can keep things low profile until I get bolder...so I put up a MARTIN HOUSE (you know - for Purple Martins) on a 15 foot steel pole. I have a radial system below it on the ground. I dont feed the pole, I have thin black wires attached at the top and coming down to feed point at the bottom (I use a remote switch to change antennas) these form inverted Ls on 160-10 meters. I had to bend the 160 L (more of a W than L)around my small yard a lot. I can run up to 700 watts...never a complaint from neighbors. I work the world on 100 watts including a lot of dx on 160. This stealth stuff is almost as much fun as working dx from a tower/beam...like the WWII resistance/spy novels I liked as a kid.

I'll let you know if I can get more overt.

I,m anonymous because of my insider position!

CC&R Survival Series Reply

by K0MU on January 13, 2001 Mail this to a friend!

The previous poster mentioned that satellite antennas are allowed. Any HF antenna that covers 10 and 15 meters is a "satellite antenna." RS12 has uplink on 15 and downlink on 10. Therefore any HF antenna / beam should be allowed. The satelite industry did us a favor!!! Is this a loophole?

CC&R Survival Series Reply

by KD5LJH on January 15, 2001 Mail this to a friend!

I appreciate all the posts about the CC&R's..... When i first started reading about it back in like sept-octish, I thought oh whats the big deall..... you know... then I started checking into it more and started understanding why its a big topic. Well, I'm finally getting my first apartment next month, and I'm pretty sure I'm not going to be able to erect anything like an antenna on top of the building for my radio working fun. So, basically I just want to thank you all that have posted on this and it has helped me with ideas and ways to get around this problem. Thanks once again guys.

73's,

Jonathan
KD5LJH

CC&R Survival Series Reply

Anonymous post on January 16, 2001 Mail this to a friend!

Maybe I'm nuts, but if I saw a development that actually had a CC&R against a VEGETABLE GARDEN, I'd go look somehwere else. Antennas or no antennas.

CC&R Survival Series Reply

by W6WO on January 16, 2001 Mail this to a friend!

It is most important to be on good terms with the HOA Board, join it if you can. I obtained approval by writing a letter to the Board explaining the benefits from emergency communications that I can provide the community in times of emergency. I mentioned in my letter that one such antenna could be treated as a special case in any community without creating a precedent for others. I compared the profile of my R6000 vertical in sq ft to a DBS dish. I stated that I would assist any homeowner who encountered problems but stopped short of saying that I would fix any problem.

They asked questions about RF health and I responded by saying my power was controlled by my FCC licence and is about the same as a lightbulb. They also asked questions about the installation and I explained my method of attachment to the apex of my roof and could quote the EIA requirements for grounding.

RE: CC&R Survival Series Reply
by KC0EAO on January 16, 2001 Mail this to a friend!
Jonathan (KD5LJH):

I'm an apartment dweller and have lived in several apartments. I think the most important thing you can do is *ask your landlord* about antennas *before* you put them up. If you explain the purpose of the antennas and amateur communication, the proposed installation procedure, and the safety measures in place regarding the antennas you might be surprised how willing they are to work with you. I have an A99 (17' tall) and a Diamond X50A (6' tall, on a 15' mast) both mounted on my balcony and the landlord doesn't have a problem with them because I asked first and did not damage the apartment installing them. 73

Dan

RE: CC&R Survival Series Reply
by AE7G on January 16, 2001 Mail this to a friend!

At one time, I had a wife who was adamant about NO ugly towers, and No ugly antennas on the house. Our house was on 1 arce with many trees in the back, where the majority of land was. One day while she was a way, I put up a 1/4 wave GP for 40 meters made out of brown wire and a SO239 (I didn't know then that a delta loop would have been better. I used a pine tree to pull the fishing line up to hold the verticle wire, and fishing line to tie the radials to other trees. The radials were 6 feet off the ground, along with the feed point. The verticle portion wasn't totally verticle, it sloped front the feed point that was about 5 feet away from the pine tree. About 18 months later, we were on the back porch watchin a sunset through and over the trees in the back, and my wife wondered what that shiny thing was hanging in the air about 30 feet away from the house (my SO239) I had tons of fun with that antenna and a SB220 amplifier. And, yes, the antenna stayed up, after the shiny "free floating" thing was observed. You had to look very carefull in the winter time when there weren't leaves on the trees to find the antenna.

I remember one time when I was single and in a crowded housing development, my neighbor asked me why I had a brick and an arrow hanging from a tree in my front yard He didn't see the brown verticle wire from the top of the tree to the ground). The arrow and brick got caught in the tree when they were thrown or shot over the tree and the fishing line got caught in the tree. I explained I had used them, and they got caught in the tree, to pull a thin wire in the tree to use for my nightly SWL.

CC&R Survival Series
by AB7RG on January 17, 2001

Reply

Mail this to a friend!

Hmm... Looks like I'll have to start on the one I was working on doing back in early October of last year after I arrived out here in Chicago. I've had to be pretty creative, but so far things are working very well. I will however be very happy to return to sunny & warm Arizona where I have much less restrictions, actually there are some where I currently live out there, but I did get around them... :^)

Anyway Bill look for an article submission by me in a few weeks. I've been so busy with computer consulting and my business on the 'net that I haven't had time to do any free-lance writing (for anyone), in the past four months. I think that I still have the framework for that article on my HD and a few of the pictures that I took of a couple of my antenna setups as well. It's not much, but it works. I'm betting that we can really learn from one another on this subject...

"Where there's a ham there's a way!" 73 Clinton AB7RG/9

RE: CC&R Survival Series Reply

Anonymous post on January 17, 2001 Mail this to a friend!

I am in a college dorm at an engineering school. Due to the fact that my dorm has a "grounded" chicken wire network in all the walls under the plaster and I cannot remove the screen at all, there is little hope of a signal getting out. I use a twin-lead J-pole (I know it isn't the best, but no where to put a mag mount) in the window frame itself. I am able to get into the close, local repeaters and on packet (2m FM) without too much trouble, but since I am chief op at the local Ham Club, I can use that equipment if necessary with all the outdoor antennas. All in all, I live in a pandora's box that eats signals and am just lucky that I can get anything out at all. The neighbors don't like the intermod through the computer speakers though, hihi. If I was able to remove the screen and get an antenna out into the tree outside of my antenna, I would have get better signal reports and my pbbs wouldn't time everyone out.

CC&R Survival Series Reply

by W6OPO on January 20, 2001 Mail this to a friend!

I bought my house about 6 years ago in a now 20 year old neighborhood (San Jose CA). CC&R's are clear and legal. No antennas, no basketball hoops no, can't work on cars in driveway, keep garage doors closed.... Then there is reality.

When I was considering the the house I plotted. Seeing the conditions in the neighborhood I could see the CC&R's were not being enforced, no gross violations but RVs are here etc.

My house is next to the end of culdesac and backs on to a freeway. That means no backyard neighbors. With a streetlight on my curb first up was a 133' dipole from the streetlight to the backyard. Used phylistrand and no insulator on the street end. It is almost invisible. That was good for 40 & 80 now what about a beam?

I bought a 40' tubular tower (used) with the rotor at the base. That with a Force12 C3-S (12' boom no traps no rotor at the top) made for the lowest profile cleanest installation one could think of. The tower went up about 2 years after the dipole and been up exactly a year and the folks are peaceful.

Bob

CC&R Survival Series Reply

by K7LA on January 21, 2001 Mail this to a friend!

Our current QTH does not have CC&R's which was a major factor in our decision to purchase the property. Our previous residence was a gated community with strict CC&R's and an overly-nosey property management firm. One easy way to beat restrictions is to operate MOBILE from your vehicle. I used to sit out in the driveway operating HF with a screwdriver-type antenna while the jerk manager just stood there raising his blood pressure. It's a great feeling knowing there's nothing they can do about it.

CC&R Survival Series Reply

by N9PSR on February 3, 2001 Mail this to a friend!

Hello,
My family and I live in graduate and family housing at Western Illinois University (WIU)and any outside antennas are strictly prohibited. So I go to my wife's minivan and operate mobile (stationary). For some silly reason the state of Illinois is very reluctant to grant a driver's license to a legally blind individual. hi hi. I usually run 100 watts from my Icom 706MkII through a hustler mobile antenna. So far I've worked 7 foreign countries and about half of the states on 40m. Since Sequoia my Leader Dog is usually with me I can get away with saying "we" while operating.hi hi. although for the most part I try to avoid using hambabble on the air. One of these days I'll get up to courage to "run with the big dogs" and operate CW. A friend from my former hometown loaned me a non J-38 antique straight key- it had been used in World War two by the Army.
Some of the reasons I like working mobile (stationary)are:
1.) My five year old daughter isn't being exposed to RF.
2.) My wife can't complain about wires being run all over the apartment.
3.) Fellow hams seem to be more willing to talk to a "Mobile" station... hi hi hi
73
de
Wayne M. Scace & Leader Dog Sequoia
n9psr@arrl.net
FISTS#4409

CC&R Survival Series Reply

by N5UOA on November 16, 2001 Mail this to a friend!

Her is one for the books as to Deeds and restrictions here where I live, Maybe we can get something done as to this fracture I have found.

IN my meighborhood there are SIX(6) differant deed systems which ALL but ONE doesnt allow antennas.

I live at the end of the street,4th house form the corner NOW the side streets, and house's behind me behind my back yard can HAVE antennas but no higher than 10 feet above the roofline. HOWEVER my street for a quarter of a block is in the DEEDED company that doesnt allow for any antennas.

When I first moved in The antenna could be spotted and got a notice to take it down which I did, and then put up another one that couldnt be seen from the front of the house.

SO FAR in 10 yrs I have not gotten a notice, to take it down. I have a six element antenna and going to get a smaller one 3 element, beam. I have been lucky im a disabled person and this is what I do to stay active as a Disabled Veteran is be a ham and help during flods and disasters. About 2 yrs ago on all the houses here on my block I have spotted satelitte antennas, and everyone I have talked to said they could put them up, SO I DONT KNOW if this blows out the ALL NO ANTENNA DEED RESTRICTION-- since they have let in one type of antenna. ALSO to my DISBELIEF is everyone has BEEN HIJACKED as to TV here, due to not being able to put up a tv antenna, we have to have cable tv installed, which to me takes away my rights to Television and putting up a antenna, WISH we could get these B******S as to HIJACKING rights.

IT is a shame that in all the areas around me people have antennas outside and this one small quarter of a blovk who is under HOUSING PROPERTY MANAGENT wont let any antennas be had and everyone else does.

LOVE DISCRIMINATION here as to what I have to live with. About a year ago I wrote a long letter to the ARRL< about how this is here to see if in the future the FCC who has NO BACKBONE anymore full of

POLITICINS turn their backs on the very HAMS who might be the only communications due to terrorism today in the US.

Thanks for reading

Mark N5UOA

CC&R Survival Series Reply
by N0VJC on February 24, 2004 Mail this to a friend!

I decided to follow the rules and apply for the antennas through normal channels. I was turned down at first and went before the board to state my case. After some muscle flexing by the board and some ego stroking by me, they approved the installation of my Hustler 4-BTV. I also use a Webster Bandspanner mounted on the ground, but this one is short enough that it doesn't fall under the rules. For two meters, I have an Isopole 144 mounted to the back of the chimney. Since it is not visible for anyplace but my backyard, it is not a problem either.
In my case, an unexpected ally turned out to be the manager of the Management Company. She explained to the board that during times of disaster, having a ham in the neighborhood is invaluable. I cannot wait for some legislation to assist hams with their antenna issues, but in the interim, I would advise hams to at least try to follow the rules and be reasonable in their request. I have no doubt that they would have turned me down completely if I had asked for "forgiveness" instead of "permission". I want to mention though, that I was going to use my 4-BTV even if I had been turned down. A wing nut at the bottom would have made it easy to remove from the ground mount, so that I could have taken it down when I wasn't using it. A pain in the neck, but better than nothing and not against the rules. Unfortunately not everyone's antennas lend themselves to this type of solution.
73,
-Steve Weidman N0VJC

CC&R Survival Series Reply
by N2CTZ on April 13, 2004 Mail this to a friend!

how about we just refuse to buy houses where there are c&r and hoa-i just found an older house on one acre no c&rs no hoa i can have antennaes ,poultry and horses!

tell all your friends not to buy houses with c&rs!

RE: CC&R Survival Series Reply
by KC4UAI on August 4, 2004 Mail this to a friend!

Not Buying a house with CCNR's is becoming more and more difficult. Most newer homes will have CCNRs, and the remaining older homes are either too expensive, or are in undesireable locations.

I checked in my local area and found that about 90% of the houses that I could afford had CCNRs. I suspect that this percentage would be less in a more rural area, but one has to live where they can earn a living.

I was wondering... Has anybody conducted a formal study to determine the % of houses that have CCNRs, verses the % that don't. Would not the same argument that the FCC used to justify state and local preemption, be valid for CCNR's if the percentage of homes availble was very small?

I would suspect that the % of available houses w/o CCNRs has been drasticly reduced since the PRB-1 rules where put into effect, and I sometimes wonder if the FCC would listen to that argument. I suppose it's wishful thinking, so we can hope that the CCNR bill makes it through congress..

73's
-= KC4UAI =-

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FEDERAL COMMUNICATIONS COMMISSION

FACT SHEET

May 2001

Over-the-Air Reception Devices Rule

Preemption of Restrictions on Placement of Direct Broadcast Satellite, Multichannel Multipoint Distribution Service, and Television Broadcast Antennas

Quick Links to Document Sections Below

- Questions and Answers
- Links to Relevant Orders and the Rule
- Guidance on Filing a Petition
- Where to Call for More Information

As directed by Congress in Section 207 of the Telecommunications Act of 1996, the Federal Communications Commission adopted the Over-the-Air Reception Devices Rule concerning governmental and nongovernmental restrictions on viewers' ability to receive video programming signals from direct broadcast satellites ("DBS"), multichannel multipoint distribution (wireless cable) providers ("MMDS"), and television broadcast stations ("TVBS").

The rule is cited as 47 C.F.R. Section 1.4000 and has been in effect since October 14, 1996. It prohibits restrictions that impair the installation, maintenance or use of antennas used to receive video programming. The rule applies to video antennas including direct-to- home satellite dishes that are less than one meter (39.37") in diameter (or of any size in Alaska), TV antennas, and wireless cable antennas. The rule prohibits most restrictions that: (1) unreasonably delay or prevent installation, maintenance or use; (2) unreasonably increase the cost of installation, maintenance or use; or (3) preclude reception of an acceptable quality signal.

Effective January 22, 1999, the Commission amended the rule so that it also applies to rental property where the renter has an exclusive use area, such as a balcony or patio.

On October 25, 2000, the Commission further amended the rule so that it applies to customer-end antennas that receive and transmit fixed wireless signals. This amendment became effective on May 25,

2001.

The rule applies to viewers who place antennas that meet size limitations on property that they own or rent and that is within their exclusive use or control, including condominium owners and cooperative owners, and tenants who have an area where they have exclusive use, such as a balcony or patio, in which to install the antenna. The rule applies to townhomes and manufactured homes, as well as to single family homes.

The rule allows local governments, community associations and landlords to enforce restrictions that do not impair the installation, maintenance or use of the types of antennas described above, as well as restrictions needed for safety or historic preservation. In addition, under some circumstances, the availability of a central or common antenna can be used by a community association or landlord to restrict the installation of individual antennas. In addition, the rule does not apply to common areas that are owned by a landlord, a community association, or jointly by condominium or cooperative owners. Such common areas may include the roof or exterior wall of a multiple dwelling unit. Therefore, restrictions on antennas installed in or on such common areas are enforceable.

This fact sheet provides general answers to questions that may arise about the implementation of the rule, but is not the rule itself. For further information or a copy of the rule, call the Federal Communications Commission at 888-CALLFCC (toll free) or (202) 418-7096. The rule is also available via the Internet by going to links to relevant Orders and the rule.

Q: What types of antennas are covered by the rule?

A: The rule applies to the following types of video antennas:

(1) A "dish" antenna that is one meter (39.37") or less in diameter (or any size dish if located in Alaska) and is designed to receive direct broadcast satellite service, including direct-to-home satellite service, or to receive or transmit fixed wireless signals via satellite.

(2) An antenna that is one meter or less in diameter or diagonal measurement and is designed to receive video programming services via MMDS (wireless cable) or to receive or transmit fixed wireless signals other than via satellite.

(3) An antenna that is designed to receive local television broadcast signals. Masts higher than 12 feet above the roofline may be subject to local permitting requirements.

In addition, antennas covered by the rule may be mounted on "masts" to reach the height needed to receive or transmit an acceptable quality signal (e.g. maintain line-of-sight contact with the transmitter or view the satellite). Masts higher than 12 feet above the roofline may be subject to local permitting requirements for safety purposes. Further, masts that extend beyond an exclusive use area may not be covered by this rule.

Q: What are "fixed wireless signals"?

A: "Fixed wireless signals" are any commercial non-broadcast communications signals transmitted via wireless technology to and/or from a fixed customer location. Examples include wireless signals used to provide telephone service or high-speed Internet access to a fixed location. This definition does **not** include, among other things, AM/FM radio, amateur ("HAM") radio, Citizens Band ("CB") radio, and Digital Audio Radio Services ("DARS") signals.

Q: Does the rule apply to hub or relay antennas?

A: The rule applies to "customer-end antennas" which are antennas placed at a customer location for the purpose of providing service to customers at that location. The rule does not cover antennas used to transmit signals to and/or receive signals from multiple customer locations.

Q: What types of restrictions are prohibited?

A: The rule prohibits restrictions that impair a person's ability to install, maintain, or use an antenna covered by the rule. The rule applies to state or local laws or regulations, including zoning, land-use or building regulations, private covenants, homeowners' association rules, condominium or cooperative association restrictions, lease restrictions, or similar restrictions on property within the exclusive use or control of the antenna user where the user has an ownership or leasehold interest in the property. A restriction impairs if it: unreasonably delays or prevents use of; (2) unreasonably increases the cost of; or (3) precludes a person from receiving or transmitting an acceptable quality signal from an antenna covered under the rule. The rule does not prohibit legitimate safety restrictions or restrictions designed to preserve designated or eligible historic or prehistoric properties, provided the restriction is no more burdensome than necessary to accomplish the safety or preservation purpose.

Q: What types of restrictions unreasonably delay or prevent viewers from using an antenna?

A: A local restriction that prohibits all antennas would prevent viewers from receiving signals, and is prohibited by the Commission's rule. Procedural requirements can also unreasonably delay installation, maintenance or use of an antenna covered by this rule. For example, local regulations that require a person to obtain a permit or approval prior to installation create unreasonable delay and are generally prohibited. Permits or prior approval necessary to serve a legitimate safety or historic preservation purpose may be permissible.

Q: What is an unreasonable expense?

A: Any requirement to pay a fee to the local authority for a permit to be allowed to install an antenna would be unreasonable because such permits are generally prohibited. It may also be unreasonable for a local government, community association or landlord to require a viewer to incur additional costs

associated with installation. Things to consider in determining the reasonableness of any costs imposed include: (1) the cost of the equipment and services, and (2) whether there are similar requirements for comparable objects, such as air conditioning units or trash receptacles. For example, restrictions cannot require that expensive landscaping screen relatively unobtrusive DBS antennas. A requirement to paint an antenna so that it blends into the background against which it is mounted would likely be acceptable, provided it will not interfere with reception or impose unreasonable costs.

Q: What restrictions prevent a viewer from receiving an acceptable quality signal?

A: For antennas designed to receive analog signals, such as TVBS, a requirement that an antenna be located where reception would be impossible or substantially degraded is prohibited by the rule. However, a regulation requiring that antennas be placed where they are not visible from the street would be permissible if this placement does not prevent reception of an acceptable quality signal or impose unreasonable expense or delay. For example, if installing an antenna in the rear of the house costs significantly more than installation on the side of the house, then such a requirement would be prohibited. If, however, installation in the rear of the house does not impose unreasonable expense or delay or preclude reception of an acceptable quality signal, then the restriction is permissible and the viewer must comply.

The acceptable quality signal standard is different for devices designed to receive digital signals, such as DBS antennas, digital MMDS antennas, digital television ("DTV") antennas, and digital fixed wireless antennas. For a digital antenna to receive or transmit an acceptable quality signal, the antenna must be installed where it has an unobstructed, direct view of the satellite or other device from which signals are received or to which signals are to be transmitted. Unlike analog antennas, digital antennas, even in the presence of sufficient over-the-air signal strength, will at times provide no picture or sound unless they are placed and oriented properly.

Q: Are all restrictions prohibited?

A: No, many restrictions are permitted. Clearly-defined, legitimate safety restrictions are permitted even if they impair installation, maintenance or use provided they are necessary to protect public safety and are no more burdensome than necessary to ensure safety. Examples of valid safety restrictions include fire codes preventing people from installing antennas on fire escapes; restrictions requiring that a person not place an antenna within a certain distance from a power line; and installation requirements that describe the proper method to secure an antenna. The safety reason for the restriction must be written in the text, preamble or legislative history of the restriction, or in a document that is readily available to antenna users, so that a person wanting to install an antenna knows what restrictions apply. Safety restrictions cannot discriminate between objects that are comparable in size and weight and pose the same or a similar safety risk as the antenna that is being restricted.

Restrictions necessary for historic preservation may also be permitted even if they impair installation, maintenance or use of the antenna. To qualify for this exemption, the property may be any prehistoric or

historic district, site, building, structure or object included in, or eligible for inclusion on, the National Register of Historic Places. In addition, restrictions necessary for historic preservation must be no more burdensome than necessary to accomplish the historic preservation goal. They must also be imposed and enforced in a non-discriminatory manner, as compared to other modern structures that are comparable in size and weight and to which local regulation would normally apply.

Q: How does the rule apply to restrictions on radiofrequency (RF) exposure from antennas that have the capability to transmit signals?

A: All transmitters regulated by the Commission, including the customer-end fixed wireless antennas (either satellite or terrestrial) covered under the amended rule, are required to meet the applicable Commission guidelines regarding RF exposure limits. The limits established in the guidelines are designed to protect the public health with a large margin of safety. These limits have been endorsed by federal health and safety agencies, such as the Environmental Protection Agency and the Food and Drug Administration. The Commission requires that providers of fixed wireless service exercise reasonable care to protect users and the public from RF exposure in excess of the Commission's limits. In addition, as a condition of invoking protection under the rule from government, landlord, and association restrictions, a provider of fixed wireless service must ensure that customer-end antennas are labeled to give notice of potential RF safety hazards posed by these antennas.

It is recommended that antennas that both receive and transmit signals be installed by professional personnel to maximize effectiveness and minimize the possibility that the antenna will be placed in a location that is likely to expose subscribers or other persons to the transmit signal at close proximity and for an extended period of time. In general, associations, landlords, local governments and other restricting entities may not require professional installation for receive-only antennas, such as one-way DBS satellite dishes. However, local governments, associations, and property owners may require professional installation for **transmitting** antennas based on the safety exception to the rule. Such safety requirements must be: (1) clearly defined; (2) based on a legitimate safety objective (such as bona fide concerns about RF radiation) which is articulated in the restriction or readily available to antenna users; (3) applied in a non-discriminatory manner; and (4) no more burdensome than necessary to achieve the articulated objectives.

For additional information about the Commission's RF exposure limits, please visit http://www.fcc.gov/oet/rfsafety or call the RF Safety Information Line at 202-418-2464.

Q: Whose antenna restrictions are prohibited?

A: The rule applies to restrictions imposed by local governments, including zoning, land-use or building regulations; by homeowner, townhome, condominium or cooperative association rules, including deed restrictions, covenants, by-laws and similar restrictions; and by manufactured housing (mobile home) park owners and landlords, including lease restrictions. The rule only applies to restrictions on property where the viewer has an ownership or leasehold interest and exclusive use or control.

Q: If I live in a condominium or an apartment building, does this rule apply to me?

A: The rule applies to antenna users who live in a multiple dwelling unit building, such as a condominium or apartment building, if the antenna user has an exclusive use area in which to install the antenna. "Exclusive use" means an area of the property that only you, and persons you permit, may enter and use to the exclusion of other residents. For example, your condominium or apartment may include a balcony, terrace, deck or patio that only you can use, and the rule applies to these areas. The rule does not apply to common areas, such as the roof, the hallways, the walkways or the exterior walls of a condominium or apartment building. Restrictions on antennas installed in these common areas are not covered by the Commission's rule. For example, the rule would **not** apply to prohibit restrictions that prevent drilling through the exterior wall of a condominium or rental unit.

Q: Does the rule apply to condominiums or apartment buildings if the antenna is installed so that it hangs over or protrudes beyond the balcony railing or patio wall?

A: No. The rule does not prohibit restrictions on antennas installed beyond the balcony or patio of a condominium or apartment unit if such installation is in, on, or over a common area. An antenna that extends out beyond the balcony or patio is usually considered to be in a common area that is not within the scope of the rule. Therefore, the rule does not apply to a condominium or rental apartment unit unless the antenna is installed wholly within the exclusive use area, such as the balcony or patio.

Q: Does the fact that management or the association has the right to enter these areas mean that the resident does not have exclusive use?

A: No. The fact that the building management or the association may enter an area for the purpose of inspection and/or repair does not mean that the resident does not have exclusive use of that area. Likewise, if the landlord or association regulates other uses of the exclusive use area (e.g., banning grills on balconies), that does not affect the viewer's rights under the Commission's rule. This rule permits persons to install antennas on property over which the person has *either* exclusive use *or* exclusive control. Note, too, that nothing in this rule changes the landlord's or association's right to regulate use of exclusive use areas for other purposes. For example, if the lease prohibits antennas and flags on balconies, only the prohibition of antennas is eliminated by this rule; flags would still be prohibited.

Q: Does the rule apply to residents of rental property?

A: Yes. Effective January 22, 1999, renters may install antennas within their leasehold, which means inside the dwelling or on outdoor areas that are part of the tenant's leased space and which are under the exclusive use or control of the tenant. Typically, for apartments, these areas include balconies, balcony railings, and terraces. For rented single family homes or manufactured homes which sit on rented property, these areas include the home itself and patios, yards, gardens or other similar areas. If renters do not have access to these outside areas, the tenant may install the antenna inside the rental unit. Renters are not required to obtain the consent of the landlord prior to installing an antenna in these areas. The rule

does not apply to common areas, such as the roof or the exterior walls of an apartment building. Generally, balconies or patios that are shared with other people or are accessible from other units are not considered to be exclusive use areas.

Q: Are there restrictions that may be placed on residents of rental property?

A: Yes. A restriction necessary to prevent damage to leased property may be reasonable. For example, tenants could be prohibited from drilling holes through exterior walls or through the roof. However, a restriction designed to prevent ordinary wear and tear (*e.g.*, marks, scratches, and minor damage to carpets, walls and draperies) would likely not be reasonable provided the antenna is installed wholly within the antenna user's own exclusive use area.

In addition, rental property is subject to the same protection and exceptions to the rule as owned property. Thus, a landlord may impose other types of restrictions that do not impair installation, maintenance or use under the rule. The landlord may also impose restrictions necessary for safety or historic preservation.

Q: If I live in a condominium, cooperative, or other type of residence where certain areas have been designated as "common," do these rules apply to me?

A: The rules apply to residents of these types of buildings, but the rules do not permit you to install an antenna on a common area, such as a walkway, hallway, community garden, exterior wall or the roof. However, you may install the antenna wholly within a balcony, deck, patio, or other area where you have exclusive use.

Drilling through an exterior wall, *e.g.* to run the cable from the patio into the unit, is generally not within the protection of the rule because the exterior wall is generally a common element. You may wish to check with your retailer or installer for advice on how to install the antenna without drilling a hole. Alternatively, your landlord or association may grant permission for you to drill such a hole. The Commission's rules generally do not cover installations if you drill through a common element.

Q: If my association, building management, landlord, or property owner provides a central antenna, may I install an individual antenna?

A: Generally, the availability of a central antenna may allow the association, landlord, property owner, or other management entity to restrict the installation by individuals of antennas otherwise protected by the rule. Restrictions based on the availability of a central antenna will generally be permissible provided that: (1) the person receives the particular video programming or fixed wireless service that the person desires and could receive with an individual antenna covered under the rule (*e.g.*, the person would be entitled to receive service from a specific provider, not simply a provider selected by the association); (2) the signal quality of transmission to and from the person's home using the central antenna is as good as, or better than, than the quality the person could receive or transmit with an individual antenna covered by

the rule; (3) the costs associated with the use of the central antenna are not greater than the costs of installation, maintenance and use of an individual antenna covered under the rule; and (4) the requirement to use the central antenna instead of an individual antenna does not unreasonably delay the viewer's ability to receive video programming or fixed wireless services.

Q: May the association, landlord, building management or property owner restrict the installation of an individual antenna because a central antenna will be available in the future?

A: It is not the intent of the Commission to deter or unreasonably delay the installation of individual antennas because a central antenna may become available. However, persons could be required to remove individual antennas once a central antenna is available if the cost of removal is paid by the landlord or association and the user is reimbursed for the value of the antenna. Further, an individual who wants video programming or fixed wireless services other than what is available through the central antenna should not be unreasonably delayed in obtaining the desired programming or services either through modifications to the central antenna, installation of an additional central antenna, or by using an individual antenna.

Q: I live in a townhome community. Am I covered by the FCC rule?

A: Yes. If you own the whole townhouse, including the walls and the roof and the land under the building, then the rule applies just as it does for a single family home, and you may be able to put the antenna on the roof, the exterior wall, the backyard or any other place that is part of what you own. If the townhouse is a condominium, then the rule applies as it does for any other type of condominium, which means it applies only where you have an exclusive use area. If it is a condominium townhouse, you probably cannot use the roof, the chimney, or the exterior walls unless the condominium association gives you permission. You may want to check your ownership documents to determine what areas are owned by you or are reserved for your exclusive use.

Q: I live in a condominium with a balcony, but I cannot receive a signal from the satellite because my balcony faces north. Can I use the roof?

A: No. The roof of a condominium is generally a common area, not an area reserved for an individual's exclusive use. If the roof is a common area, you may not use it unless the condominium association gives you permission. The condominium is not obligated to provide a place for you to install an antenna if you do not have an exclusive use area.

Q: I live in a mobile home that I own but it is located in a park where I rent the lot. Am I covered by the FCC rule?

A: Yes. The rule applies if you install the antenna anywhere on the mobile or manufactured home that is owned by you. The rule also applies to antennas installed on the lot or pad that you rent, as well as to other areas that are under your exclusive use and control. However, the rule does not apply if you want to

install the antenna in a common area or other area outside of what you rent.

Q: I want a conventional "stick" antenna to receive a distant over-the air television signal. Does the rule apply to me?

A: No. The rule does not apply to television antennas used to receive a distant signal.

Q: I want to install an antenna for broadcast radio or amateur radio. Does the rule apply to me?

A: No. The rule does not apply to antennas used for AM/FM radio, amateur ("ham") radio, Citizen's Band ("CB") radio or Digital Audio Radio Services ("DARS").

Q: I want to install an antenna to access the Internet. Does the rule apply to me?

A: Yes. Antennas designed to receive and/or transmit data services, including Internet access, are included in the rule.

Q: Does this mean that I can install an antenna that will be used for voice and data services even though it does not provide video transmissions?

A: Yes. The most recent amendment expands the rule and permits you to install an antenna that will be used to transmit and/or receive voice and data services, except as noted above. The rule will also continue to cover antennas used to receive video programming.

Q: I have already installed an antenna that is used solely for the purpose of receiving video programming. Am I affected by this amendment?

A: Persons who have already installed, or who plan to install, an antenna designed to receive only video programming are not affected by this amendment. The purpose of the amendment is to permit persons to install antennas that may be used for voice and data services, as well as for video programming services. The rules concerning restrictions on the placement of video antennas will apply equally to antennas that are used for voice and data services.

Q: I'm a board member of a homeowners' association, and we want to revise our restrictions so that they will comply with the FCC rule. Do you have guidelines you can send me?

A: We do not have sample guidelines because every community is different. We can send you the rule and the relvant orders, which will give you general guidance. (See list of documents at the end of this factsheet. Some communities have written restrictions that provide a prioritized list of placement preferences so that residents can see where the association wants them to install the antenna. The residents should comply with the placement preferences provided the preferred placement does not

impose unreasonable delay or expense or preclude reception of an acceptable quality signal.

Q: What restrictions are permitted if the antenna must be on a very tall mast to get a signal?

A: If you have an exclusive use area that is covered by the rule and need to put your antenna on a mast, the local government, community association or landlord may require you to apply for a permit for safety reasons if the mast extends more than 12 feet above the roofline. If you meet the safety requirements, the permit should be granted. Note that the Commission's rule only applies to antennas and masts installed wholly within the antenna user's exclusive use area. Masts that extend beyond the exclusive use area are outside the scope of the rule. For installations on single family homes, the "exclusive use area" generally would be anywhere on the home or lot and the mast height provision is usually most relevant in these situations. For example, if a homeowner needs to install an antenna on a mast that is more than 12 feet taller than the roof of the home, the homeowners' association or local zoning authority may require a permit to ensure the safety of such an installation, but may not prohibit the installation unless there is no way to install it safely. On the other hand, if the owner of a condominium in a building with multiple dwelling units needs to put the antenna on a mast that extends beyond the balcony boundaries, such installation would generally be outside the scope and protection of the rule, and the condominium association may impose any restrictions it wishes (including an outright prohibition) because the Commission rule does not apply in this situation.

Q: Does the rule apply to commercial property or only residential property?

A: Nothing in the rule excludes antennas installed on commercial property. The rule applies to property used for commercial purposes in the same way it applies to residential property.

Q: What can a local government, association, or consumer do if there is a dispute over whether a particular restriction is valid?

A: Restrictions that impair installation, maintenance or use of the antennas covered by the rule are preempted (unenforceable) unless they are no more burdensome than necessary for the articulated legitimate safety purpose or for preservation of a designated or eligible historic site or district. If a person believes a restriction is preempted, but the local government, community association, or landlord disagrees, either the person or the restricting entity may file a Petition for Declaratory Ruling with the FCC or a court of competent jurisdiction. We encourage parties to attempt to resolve disputes prior to filing a petition. Often calling the FCC for information about how the rule works and applies in a particular situation can help to resolve the dispute. If a local government, community association, or landlord acknowledges that its restriction impairs installation, maintenance, or use and is preempted under the rule but believes it can demonstrate "highly specialized or unusual" concerns, the restricting entity may apply to the Commission for a waiver of the rule.

Q: What is the procedure for filing a petition or requesting a waiver at the Commission?

A: There is no special form for a petition. You may simply describe the facts, including the specific restriction(s) that you wish to challenge. If possible, attach a copy of the restriction(s) and any relevant correspondence. If this is not possible, be sure to include the exact language of the restriction in question with the petition. General or hypothetical questions about the application or interpretation of the rule cannot be accepted as petitions.

Petitions for declaratory rulings and waivers must be served on all interested parties. For example, if a homeowners' association files a petition seeking a declaratory ruling that its restriction is not preempted and is seeking to enforce the restriction against a specific resident, service must be made on that specific resident. The homeowners' association will not be required to serve all other members of the association, but must provide reasonable, constructive notice of the proceeding to other residents whose interests foreseeably may be affected. This may be accomplished, for example, by placing notices in residents' mailboxes, by placing a notice on a community bulletin board, or by placing the notice in an association newsletter. If a local government seeks a declaratory ruling or a waiver from the Commission, the local government must take steps to afford reasonable, constructive notice to residents in its jurisdiction (*e.g.*, by placing a notice in a local newspaper of general circulation). Proof of constructive notice must be provided with a petition. In this regard, the petitioner should provide a copy of the notice and an explanation of where the notice was placed and how many people the notice reasonably might have reached.

Finally, if a person files a petition or lawsuit challenging a local government's ordinance, an association's restriction, or a landlord's lease, the person must serve the local government, association or landlord, as appropriate. You must include a "proof of service" with your petition. Generally, the "proof of service" is a statement indicating that on the same day that your petition was sent to the Commission, you provided a copy of your petition (and any attachments) to the person or entity that is seeking to enforce the antenna restriction. The proof of service should give the name and address of the parties served, the date served, and the method of service used (*e.g.*, regular mail, personal service, certified mail).

All allegations of fact contained in petitions and related pleadings before the Commission must be supported by an affidavit signed by one or more persons who have actual knowledge of such facts. You must send an original and two copies of the petition and all attachments to: Secretary, Federal Communications Commission, 445 12th Street, S.W., Washington, D.C. 20554, Attention: Media Bureau.

Q: Can I continue to use my antenna while the petition or waiver request is pending?

A: Yes, unless the restriction being challenged or for which a waiver is sought is necessary for reasons of safety or historic preservation. Otherwise, the restriction cannot be enforced while the petition is pending.

Q: Who is responsible for showing that a restriction is enforceable?

A: When a conflict arises about whether a restriction is valid, the local government, community association, property owner, or management entity that is trying to enforce the restriction has the burden of proving that the restriction is valid. This means that no matter who questions the validity of the restriction, the burden will always be on the entity seeking to enforce the restriction to prove that the restriction is permitted under the rule or that it qualifies for a waiver.

Q: Can I be fined and required to remove my antenna immediately if the Commission determines that a restriction is valid?

A: If the Commission determines that the restriction is valid, you will have a minimum of 21 days to comply with this ruling. If you remove your antenna during this period, in most cases you cannot be fined. However, this 21-day grace period does not apply if the FCC rule does not apply to your installation (for example, if the antenna is installed on a condominium general common element or hanging outside beyond an apartment balcony. If the FCC rule does not apply at all in your case, the 21-day grace period does not apply.

Q: Who do I call if my town, community association or landlord is enforcing an invalid restriction?

A: Call the Federal Communications Commission at (888) CALLFCC (888-225-5322), which is a toll-free number, or 202-418-7096, which is not toll-free. Some assistance may also be available from the direct broadcast satellite company, multichannel multipoint distribution service, television broadcast station, or fixed wireless company whose service is desired.

Links to Relevant Orders and the Rule

- (First) Report and Order, FCC 96-328, released August 6, 1996: [Text Version | WordPerfect Version]
- Declaratory Ruling, Star Lambert, DA 97-1554, released July 27, 1997: [Text]
- Declaratory Ruling, Jay Lubliner, DA 97-2188, released October 14, 1997: [Text]
- Declaratory Ruling, Michael MacDonald, DA 97-2189, released October 14, 1997: [Text]
- Declaratory Ruling, Omnivision, DA 97-2187, released October 14, 1997: [Text]
- Declaratory Ruling, Wireless Broadcasting Systems (WBSS), DA 97-2506, released November 28, 1997: [WordPerfect | Text]
- Declaratory Ruling, Victor Frankfurt, DA 97-2305, released December 31, 1997: [Text]
- Declaratory Ruling, Jason Peterson, DA 98-0188, released February 4, 1998: [Text]
- Declaratory Ruling, Jordan Lourie, DA 98-1170, released June 17, 1998: [WordPerfect | Text]
- Declaratory Ruling, James Sadler, DA 98-1284, released July 1, 1998: [WordPerfect | Text]
- Memorandum Opinion and Order, Denial of Application of Review of Declaratory Ruling for Jay Lubliner (above), FCC 98-201, released August 21, 1998: [WordPerfect | Text]

- Order on Reconsideration, FCC 98-214, released September 25, 1998: [WordPerfect | Text]
- Second Report and Order, FCC 98-273, released November 20, 1998: [Text | WordPerfect | Acrobat | News Release and Statements]
- Declaratory Ruling, Stanley and Vera Holliday, DA 99-2132, released October 8, 1999: [MSWord | Acrobat]
- Second Order on Reconsideration, FCC 99-360, released November 24, 1999: [Text | MSWord]
- Declaratory Ruling, Bell Atlantic Video, DA 00-927, released April 26, 2000: [MSWord | Acrobat]
- Competitive Networks Report and Order, FCC 00-366, released October 25, 2000: [Text | MSWord | Acrobat | News Release and Statements]
- Declaratory Ruling, Victor Frankfurt, DA 01-0153, released February 7, 2001: [MSWord | Acrobat]
- Declaratory Ruling, Corey Roberts, DA 01-1276, released May 24, 2001: [MSWord | Acrobat]
- Memorandum Opinion and Order, Denial of Application of Review of Declaratory Ruling for Victor Frankfurt (above), FCC 03-210, released August 27, 2003 : [MSWord | Acrobat]
- Memorandum Opinion and Order, Philip Wojcikewicz, DA 03-2971, released September 29, 2003: [MSWord | Acrobat]
- OTARD Rule, 47 C.F.R. Section 1.4000.

GUIDANCE ON FILING A PETITION

Q: What are the procedural requirements for filing a Petition for Declaratory Ruling or Waiver with the Commission?

A: There is no special form for a petition. You may simply describe the facts, including the specific restriction(s) that you wish to challenge. If possible, attach a copy of the restriction(s) and any relevant correspondence. If this is not possible, be sure to include the exact language of the restriction in question with the petition. General or hypothetical questions about the application or interpretation of the rule cannot be accepted as petitions.

Petitions for declaratory rulings and waivers must be served on all interested parties. An entity seeking to impose or maintain a restriction must include with its petition a proof of service that it has served the affected residents. Similarly, an antenna user seeking to challenge the permissibility of a restriction must include with the petition a proof of service that the antenna user has served the restricting entity with a copy of the Petition.

If you are an antenna user, you must serve a copy of the Petition on the entity seeking to enforce the restriction (*i.e.*, the local government, community association or landlord). If you are a local government, community association or landlord, you must serve a copy of the Petition on the residents in the

community who currently have or wish to install antennas that will be affected by the restriction your Petition seeks to maintain. For example, if a homeowners' association files a petition seeking a declaratory ruling that its restriction is not preempted and is seeking to enforce the restriction against a specific resident, service must be made on that specific resident. The homeowners' association will not be required to serve all other members of the association, but must provide reasonable, constructive notice of the proceeding to other residents whose interests may foreseeably be affected. This may be accomplished, for example, by placing notices in residents' mailboxes, by placing a notice on a community bulletin board, or by placing the notice in an association newsletter. If a local government seeks a declaratory ruling or a waiver from the Commission, the local government must take steps to afford reasonable, constructive notice to residents in its jurisdiction (*e.g.*, by placing a notice in a local newspaper of general circulation). Proof of constructive notice must be provided with a petition. In this regard, the petitioner should provide a copy of the notice and an explanation of where the notice was placed and how many people the notice might reasonably have reached.

Finally, if a person files a petition or lawsuit challenging a local government's ordinance, an association's restriction, or a landlord's lease, the person must serve the local government, association or landlord, as appropriate. You must include a "proof of service" with your petition. Generally, the "proof of service" is a statement indicating that on the same day that your petition was sent to the Commission, you provided a copy of your petition (and any attachments) to the person or entity that is seeking to enforce the antenna restriction. The proof of service should give the name and address of the parties served, the date served, and the method of service used (*e.g.*, regular mail, personal service, certified mail).

If you wish to file either a Petition for Declaratory Ruling or a Petition for Waiver pursuant to the Commission's Over-the-Air Reception Devices Rule (47 CFR Section 1.4000), you must file an original and two copies of your Petition on the following address:

Office of the Secretary
Federal Communications Commission
445 12th Street, S.W.
Washington, D.C. 20554
Attn: Media Bureau

Q: What are the substantive requirements for filing a petition for waiver or declaratory ruling?

A: To file a Petition for Waiver, follow the requirements in Section 1.4000(c) of the rule. The local government, community association or landlord requesting the waiver must demonstrate "local concerns of a highly specialized or unusual nature." The petition must also specify the restriction for which the waiver is sought, or the petition will not be considered.

To file a Petition for Declaratory Ruling, follow the requirements set forth in Section 1.4000(d) of the rule. Set out the restriction in question so that we can determine whether it is permissible or prohibited under the rule. In a Petition for Declaratory Ruling, the burden of demonstrating that a particular

restriction complies with the rule is on the entity seeking to impose the restriction (*e.g.*, the local government, community association or landlord).

We recommend that you include the language of the restriction in question, as well as a daytime telephone number, with your petition.

While a petition for declaratory ruling or waiver is pending with the Commission or a court, the restriction in question may not be enforced unless it is necessary for safety or historic preservation. No fines or penalties, including attorneys fees, may be imposed by the restricting entity while a petition is pending. If the restriction is found to be permissible, the antenna users subject to the ruling will generally have at least 21 days in which to comply before a fine or penalty is imposed.

- FCC -

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Antenna Restrictions, Power Restrictions, RFI Problems, or Limited Space keeping you off the air?

There are SOLUTIONS. See how others do it! Here are some helpful LINKS. Build your own hidden stealth and secret antennas

Learn how to build, hide and operate indoor and outdoor Ham antennas without causing RFI problems. Some of these designs were taken from pirate operations and applied to legal radio. I do not condone pirate radio, but why not apply their hideing techniques with proven ham antenna designs and theory.

---

*Newsletter archives dealing with condo and apartment Radio Station invisible antennas, interference solutions and more.

The CONDO COMMUNICATOR

---

*News Reflector dealing with low power, super efficient concealed ham antennas, solving the impossible, and having fun doing it. Their designs work with 100w stuff too!

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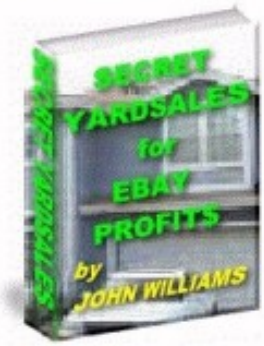
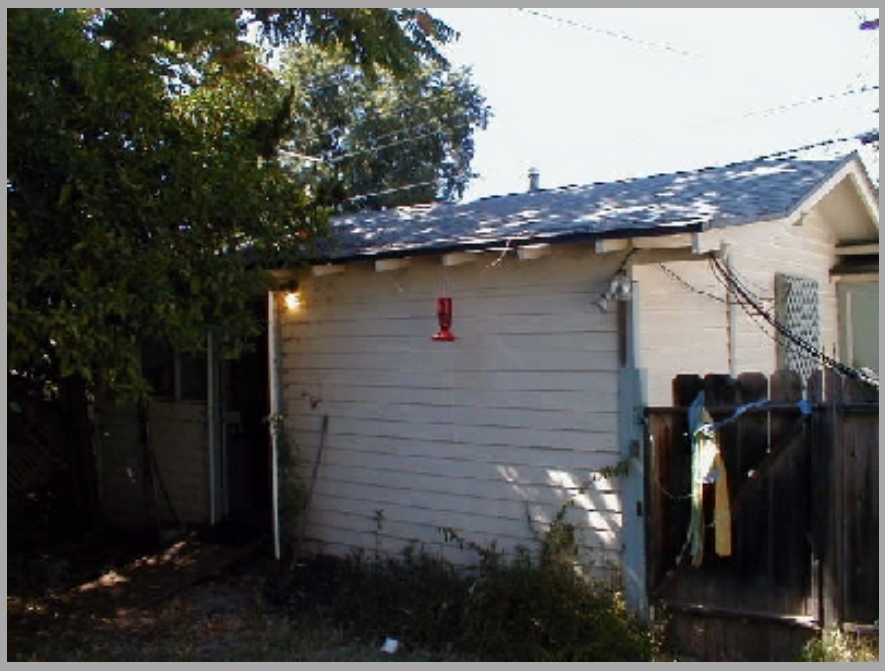


Hide your antenna in something else. Extreme Measure for the most severely restricted environment. When NOBODY can know you have a transmitter. For apartment Balcony, Great working! Hanging "HUMMINGBIRD FEEDER" bird feeder 10-40m HF Antenna Instructions. This hides the outside loading coil so your random wire will tune up on 10-40m. Puts substansial portion of your radiated RF outside your apartment-condo. Here is a brief description until I put up a page: Buy a plastic hummingbird feeder with standard glass syrup jar. Replace syrup bottle with a 2 liter soda bottle. Threads of soda bottle match feeder bottom perfectly, plus plenty of room to put a large COIL inside. Cut bottom 5" of bottle off. Save it to slide back over remainder later after you instal a 80-100 turn air or ceramic core coil. No plastic cores. Can get hot! on some freqs. Coil turn spacing min= space of same wire diameter as gap between turns. Larger gaps ok. Solder 10',(or more), wire to top of coil, run through bottle end to hang. Use a foot or two to hang and run remainder painted, hidden behind eves etc. horizontally for "INVERTED L" capacity hat. Solder a random wire to bottom of coil to exit out of bottom of feeder to your tuner (tune with tuner as a random wire. use a counterpoise, a real or artificial ground.) Small guage Black insulated wire is hardest to see. Best to hide most of feed wire, but do not run parallel next to other parallel metal objects,(ok to cross metal. just don't run along lengths of metal rails, sliding doors, extension cords, etc.). If ask what is "THAT" wire for, say, "It is a special ultrasonic LARGE bird repeller" and it only repells one foot from my feeder. Tell them how safe it is. Make up details, go on and on about technical details etc. They are sorry they ask and leave. This always works for me. Paint soda bottle fire engine red to hide coil and match rest of feeder. The longer the feed

wire the better. The more turns and diameter of coil the better. You want to increase your chances of a "PEAK CURRENT POINT" occuring somewhere, (that is why so many turns), in the coil for lowest angle radaition possible under such compromised conditions. (sorry no more syrup). Safe for birds. Works great and even the most snoopy manager neighbor will never suspect a pirate like antenna. This description and photos will get you started. You can easily improvise until I post more construction photo details.


10-40m Hummingbird feeder antenna in use

LOOK AT THESE 10-30m and 40m balcony antennas. 10-30m is an AEA Isoloop magnetic loop, (MFJ type loop), and a 40m hamstick bundled with fishing poles to hide. Stick is tuned with wire tuner and artificial ground counter poise. Loop is covered with plastic tarp to hide. I get average 559 sig reports. Loop works DX on CW! I can't say enough about these loops. They really perform. Hamstick equals a 100w moblie for performance. It,s fun and I get to keep my hobby where not allowed.

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Previous: *Surfin':* HF Propagation Software From Your Uncle Sam

Next: FCC May Be Able to Resume Vanity Processing Soon

ARRL Offers On-Line "Antenna Restrictions 'How To' Chart"

NEWINGTON, CT, Jan 7, 2002--The ARRL Regulatory Information Branch has made available what might be called a "triage center" for amateurs facing the prospect of dealing with various roadblocks to erecting an antenna system at their residence. The new "Antenna Restrictions 'How To' Chart" page offers three separate outlines that help the user to logically work through issues involving local government zoning restrictions; deed covenants, conditions and restrictions (CC&Rs); or rental/lease restrictions relating to antenna structures.

The prime focus is on dealing with local zoning restrictions to putting up an antenna structure and how to make the best possible case at a local regulatory board hearing. Some of the advice there also applies to CC&Rs and rental/lease situations too.

"Remember: at the hearing, your presentation will be 80% of the battle, and 100% of the basis for any record, if the case ends up going to court," the "Important Notice" on the page advises.

Each "how to" outline is structured around a series of questions--much like a logic or flow chart. Depending on the answer, the user is referred to specific information or additional resources. The page also offers some step-by-step suggestions. For example, the local government zoning outline suggests 10 steps to those seeking to change an overly restrictive local amateur antenna ordinance. One of them is to obtain the ARRL book *Antenna Zoning for the Radio Amateur* ($49.95; order Item 8217 from ARRL via the Online Store or order by calling toll-free

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Practical Digital Signal Processing -- **Now Shipping!** -- Gain a working knowledge of DSP applications, including

888-277-5289), which offers detailed information on working with local governments. Written by attorney Fred Hopengarten, K1VR, the book describes proven techniques and strategies that amateurs can employ in efforts to obtain an antenna-structure permit.

As the page emphasizes up front, however, neither this book nor the outlines on the "Antenna Restrictions 'How To' Chart" page are intended as substitutes for advice from an attorney.

communications and process control.

ARRL's HF Digital Handbook -- **Now Shipping!** -- 3rd edition. Use your computer to talk to the world!

The local government zoning outline also asks, "Has your state codified PRB-1?" and, if the answer is no, offers suggestions and information for amateurs who would like their legislature to incorporate the limited federal preemption known as PRB-1 into their state's laws. So far, only 13 states have done so, but bills are still in the works in a few others.

The page offers limited guidance to those confronting CC&Rs--an issue facing more and more amateurs these days--and to those who rent or lease their homes and still want to be able to install an antenna--a situation where PRB-1 does not apply. In both instances, affected amateurs are advised to develop and present logical and persuasive cases for being allowed to install an antenna system--much as they would have to do when dealing with a local government.

Previous: *Surfin':* HF Propagation Software From Your Uncle Sam

Next: FCC May Be Able to Resume Vanity Processing Soon

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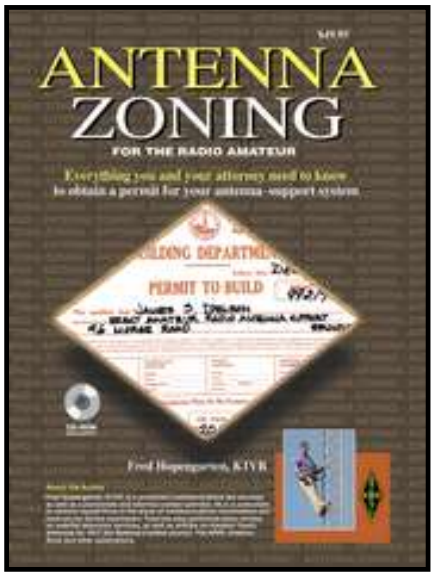
Antenna Zoning for the Radio Amateur

-- By Fred Hopengarten, K1VR



Everything you and your attorney need to know to obtain a permit for your antenna-support system.

Don't let the confusing tangle of ordinances and by-laws keep you from installing the antenna you need in order to communicate effectively! In recent years, many cities and towns have enacted ordinances designed mainly to regulate cellular antenna structures. Unfortunately, hams have sometimes been caught in the backlash of regulatory overkill. This book describes proven techniques and strategies that a ham and his or her attorney can use to obtain an antenna-structure permit (also includes material on Canadian law and regulation).

Contents

- Introduction: Why This Book Was Written
- 1: Principles That Will Help You Win
- 2: The Process in a Nutshell
- 3: Your Winning Team
- 4: Basic Preparations
- 5: Getting to Know the Players
- 6: Possible Objections
- 7: Preparing the Permit Application
- 8: Public Hearings—Your Big Moment in the Spotlight
- 9: Deliberations and Decisions
- 10: Now, Get the Permit and Build Your Antenna
- 11: Awkward Post-Permit Situations

- 12: Appeals
- 13: Tower and Antenna Regulation in Canada [by Tim Ellam, VE6SH]
- 14: Bibliography
- Appendix A: The Law
- Appendix B: Drafting a Bylaw, or Redrafting a Bylaw
- Appendix C: Lawsuits

CD-ROM included containing case law, customizable forms and additional legal reference material. Windows and Macintosh compatible.

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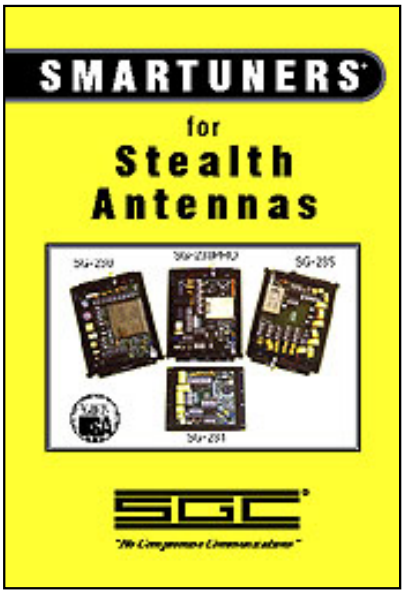

ARE YOU INSTALLING A NEW ANTENNA SYSTEM?
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A book about theory and installation of effective HF antenna systems. For novice to experienced HF users, "Smartuners for Stealth Antennas" shows you antenna installations tried and proven using SGC Smartuners. Dozens of illustrations actually show you how to set up your very own practical *and* "hidden" HF antenna system.

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Antenna Restrictions, Power Restrictions, RFI Problems, or Limited Space keeping you off the air?

There are SOLUTIONS. See how others do it! Here are some helpful LINKS. Build your own hidden stealth and secret antennas

Learn how to build, hide and operate indoor and outdoor Ham antennas without causing RFI problems. Some of these designs were taken from pirate operations and applied to legal radio. I do not condone pirate radio, but why not apply their hideing techniques with proven ham antenna designs and theory.

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*Newsletter archives dealing with condo and apartment Radio Station invisible antennas, interference solutions and more.

The CONDO COMMUNICATOR

---

*News Reflector dealing with low power, super efficient concealed ham antennas, solving the impossible, and having fun doing it. Their designs work with 100w stuff too!

QRP-L REFLECTOR updated hourly

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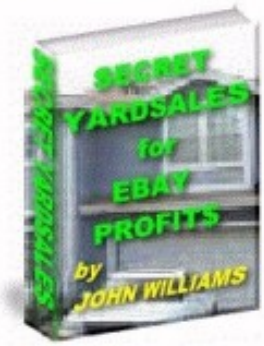
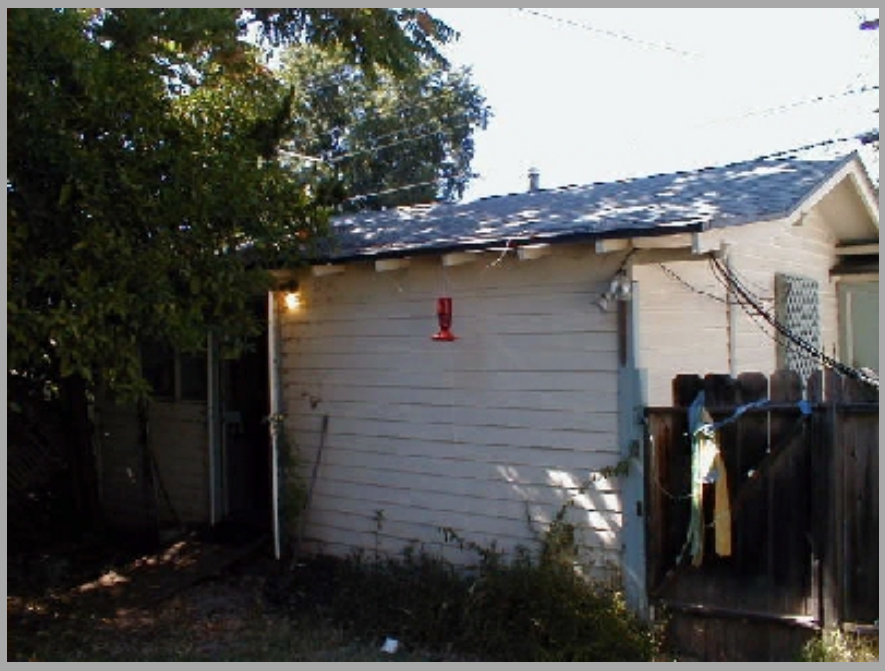


Hide your antenna in something else. Extreme Measure for the most severely restricted environment. When NOBODY can know you have a transmitter. For apartment Balcony, Great working! Hanging "HUMMINGBIRD FEEDER" bird feeder 10-40m HF Antenna Instructions. This hides the outside loading coil so your random wire will tune up on 10-40m. Puts substansial portion of your radiated RF outside your apartment-condo. Here is a brief description until I put up a page: Buy a plastic hummingbird feeder with standard glass syrup jar. Replace syrup bottle with a 2 liter soda bottle. Threads of soda bottle match feeder bottom perfectly, plus plenty of room to put a large COIL inside. Cut bottom 5" of bottle off. Save it to slide back over remainder later after you instal a 80-100 turn air or ceramic core coil. No plastic cores. Can get hot! on some freqs. Coil turn spacing min= space of same wire diameter as gap between turns. Larger gaps ok. Solder 10',(or more), wire to top of coil, run through bottle end to hang. Use a foot or two to hang and run remainder painted, hidden behind eves etc. horizontally for "INVERTED L" capacity hat. Solder a random wire to bottom of coil to exit out of bottom of feeder to your tuner (tune with tuner as a random wire. use a counterpoise, a real or artificial ground.) Small guage Black insulated wire is hardest to see. Best to hide most of feed wire, but do not run parallel next to other parallel metal objects,(ok to cross metal. just don't run along lengths of metal rails, sliding doors, extension cords, etc.). If ask what is "THAT" wire for, say, "It is a special ultrasonic LARGE bird repeller" and it only repells one foot from my feeder. Tell them how safe it is. Make up details, go on and on about technical details etc. They are sorry they ask and leave. This always works for me. Paint soda bottle fire engine red to hide coil and match rest of feeder. The longer the feed

wire the better. The more turns and diameter of coil the better. You want to increase your chances of a "PEAK CURRENT POINT" occuring somewhere, (that is why so many turns), in the coil for lowest angle radaition possible under such compromised conditions. (sorry no more syrup). Safe for birds. Works great and even the most snoopy manager neighbor will never suspect a pirate like antenna. This description and photos will get you started. You can easily improvise until I post more construction photo details.


10-40m Hummingbird feeder antenna in use

LOOK AT THESE 10-30m and 40m balcony antennas. 10-30m is an AEA Isoloop magnetic loop, (MFJ type loop), and a 40m hamstick bundled with fishing poles to hide. Stick is tuned with wire tuner and artificial ground counter poise. Loop is covered with plastic tarp to hide. I get average 559 sig reports. Loop works DX on CW! I can't say enough about these loops. They really perform. Hamstick equals a 100w moblie for performance. It,s fun and I get to keep my hobby where not allowed.

Invisible Amateur Antennas-- antenna special on hard-core- dx.com Plans, Plans and more Plans. These guys left nothing out!!!

Hard-Core-DX Hidden antennas

AMATEUR RADIO ANTENNA PROJECTS Links to many antenna designs, radio kits, designs, concealed station ideas.

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I work as a Building Engineer for an aerospace company that specializes in antenna systems. I hold an Extra Class License

N0HC's Ham Radio & More

Low Profile Operating & High Profile Fun

Hi, my name is Cliff

A lot of Hams live in areas that do not allow visible outside antennas like these in the back ground. This page is intended to provide some ideas for low profile operation.

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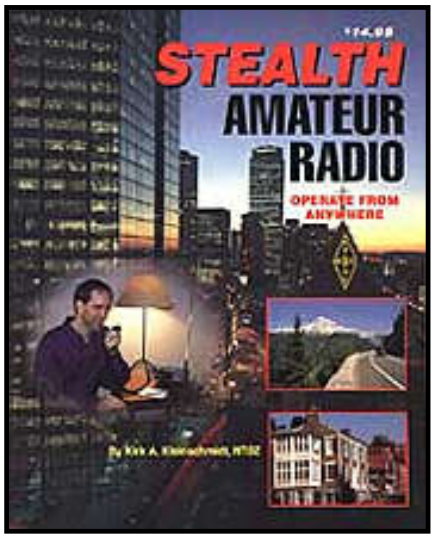

Adventure into the world of hidden stations and invisible antennas! Set up and operate a station without calling attention to yourself. Successfully operate a low power (QRP) station. Install safe antennas, including indoor antennas. Build invisible antennas. Install and operate a mobile station, to "get away" from radio-forbidden locations. Operate a portable station from a campground, motel room, picnic area, mountaintop or other location. Handle interference from your station to nearby consumer electronics devices as well as to your station from other nearby devices. Use this book and enjoy operating from just about anywhere!

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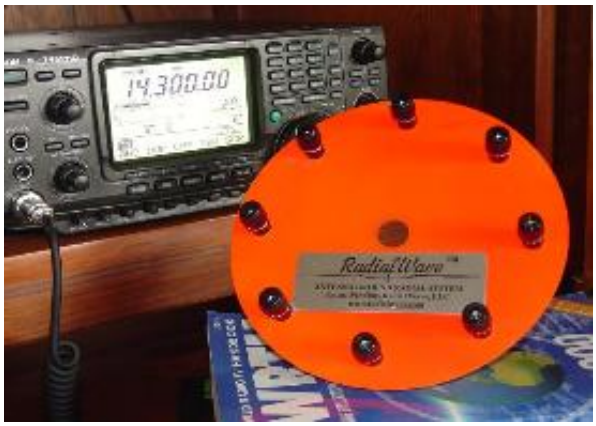

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	Saratoga EZ PSK for ICOM 703/706 	$45	More Information and Ordering
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MAG-M4	4" dia Magnetic Mount 3/8 *24	$16	More Information and Ordering

MAG-M5	5" dia Magnetic Mount 3/8 *24	$16	More Information and Ordering
SM-M	RadialWave Heavy-Duty Surface Mount (SO-239)	$5	More Information and Ordering
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Clif's Ham Radio Page

Welcome to Clif's ham radio connection. On this page you'll find links to other interesting sites. It is not my intention to create the end all page with links to everywhere but I've included a few of the ones that I think are most pertinent.

My particular area of interest is in high frequency (HF) antennas in less than optimal conditions. Most of the available programs used to analyze antennas today assume that the antenna is mounted in free space or mounted a uniform distance above ground - conditions that unfortunately many hams can get nowhere near achieving. Since the software tools are somewhat limited in what can be simulated I'm instead looking for the apartment dwellers, condo owners and other restricted antenna users with any experience. Your experience level is not important - what you did and what the results were are more important. It doesn't matter if what you tried didn't work, just let the rest of us know so we might learn from your experience. I'm collecting data - not to analyze but rather to show empirically what works and what doesn't. Send me all your details, in particular try to provide the following information:

- description of the antenna
- where and how it was mounted such as in the attic, under the eaves, in the middle of a group of trees etc.
- Typical transmitter power
- Contacts made
- Anything else you think might help another ham

I'll make the data available here as case studies. I've included the results of my own experiences to give you an idea. Got comments or questions? Ideas of how to make this site better? Email me!
As more data becomes available, the format of this page may change, however it will remain the entry point into the database with the same name and location so go ahead and bookmark it.

Summary of experience with restricted space antennas

First and foremost hams faced with restrictive antenna constraints should not lose faith. You still can operate sucessfully although you will never be able to out-slug the hams running linears into beams mounted high up on towers. If you build a hidden outdoor antenna you will probably be limited to 100 watts or less due to the wire sizes and insulator types. If the antenna is mounted indoors, your power will still be restricted due to health hazards to yourself or other family members (arcing or RF energy) or due to interference with televisions, computers, VCRs, telephones, etc. However the QRP hams continue to show us that QSOs can and do happen with regularity at low power outputs. In general you should strive to build an antenna as high and as close to full size as your situation allows. If elements need to be bent to fit the space thats ok - just try not to have them double back on themselves. An antenna tuner is a must have item with almost any antenna you will build. Almost anything metal can be made to radiate RF

energy so don't despair! If you meet with sucess, drop me some email so we can share it with others. Some other common approaches are the flagpole vertical, the magnetic loop and slinky dipole. Another approach I want to try is to build a St. Louis Vertical which is a portable antenna that works 10m to 40m and only erect it when I'm on the air.

Case 1: 10m dipole in attic

Antenna type

Full size dipole made with #24 gauge magnet wire and ceramic insulators. Oriented approximately NNW by SSE. Actual elevation above ground was approximately 25 to 30 feet.

Structure

Two and a half story condominium unit. Shack was in upper level with the antenna mounted in the attic using cord to mount the antenna between rafters. Structure was wood frame with a stucco exterior and a composite shingle roof. Also under one half of the dipole was a forced air furnace made of galvanized steel. This heater was about 2.5 feet below the wire of the dipole. Shack was located in Southern California.

Rig Details

Icom IC725 running average of 50 watts. Coupled to antenna through RG58/U coax. Used MFJ-941E tuner to match rig to antenna.

Performance Summary

When the 10m band opened, 50 watts and my attic dipole did quite well. Most signal reports were 55 to 59. Contacts were made to Indiana, Alabama, Louisiana, Texas, Wisconsin, Kansas, Minnesota, Mississippi, Pennslyvania, North Carolina, New York, Rhode Island, Hawaii plus a few other states and even Japan!

Comments

This antenna may have done well since it was in a wood structure and the actual antenna height above ground was greater than a half wavelength. The contacts appeared consistent with the orientation of the antenna, that is most of the contact occured broadside to the antenna.

Case 2: 40m twin lead Marconi in attic

Antenna type

This antenna was a twin lead Marconi based on information found in William Orr's **Radio Handbook.** This antenna was also fed with RG-58/U coax. In addition, a half wavelength counterpoise was attached to the coax shield/antenna junction to provide an RF ground.

Structure

This antenna was also mounted in the same attic as case 1. The twin lead was snaked around the perimeter of the attic and held in place by attaching it to the studs with cable ties. The twin lead could not be laid out in a straight line instead it was routed in a shape like a question mark where the bottom of the question mark was the feed point. The first section was about 20 feet followed by a 90 degree bend and the next 10 feet followed by another 90 degree bend and the final 3 or 4 feet. The first section was oriented roughly E-W, the second section N-S and the third section was

again E-W. The counterpoise was snaked in the other direction in a similar shape with the first section of about 25' going N-S, and the remainder going E-W.

Rig Details

Same as Case 1.

Performance Summary

As expected, this antenna didn't perform like a free space 1/4 wave vertical. It took time but the MFJ tuner would get a good match to the transmitter. Normal power used was 50 watts. Contacts ranged all around the western United States with the majority in Oregon, Washington and Northern California. There were also contacts made in Idaho, Utah, Montana, Colorado, Nevada, Alaska, British Columbia and Texas.

Comments

This antenna was used to primarily work 40m CW. On occasion it was used of 15m CW also. Signal reports were average with R's of 4 to 5, S's from 3 to 9 and T's of 9. There were very few reports of 599. The signal was getting out but not very well. Almost all the contacts were broadside to the longest section of the twinlead.

Case 3: Triband inverted V dipole under the eaves

Antenna type

This antenna was constructed as a 40m dipole and a 10m dipole with a common feedpoint. The 10m elements are built using #24 gauge magnet wire and ceramic insulators. The 40m elements are built using #18 gauge hookup wire. The apex of the inverted V is at the high point of the roofline, approximately 18 to 19' above ground. The 10m elements slant downward about 30 degrees below horizontal and remain spaced about 14 inches from the stucco surface of the wall. The 40m elements are secured under the eaves and extend about 30 feet from the apex before reaching the corners of the house where they both make 90 degree bends and continue under the eaves. At this point, the elements are only about 8 feet above ground. The major portion of both antennas are oriented N-S.

Structure

The house is a wood frame, stucco exterior, single story dwelling with a concrete tile roof. The eaves extend about 16 inches beyond the walls of the house. The house is located in Southern California.

Rig Details

Same as Case 1.

Performance Summary

So far it appears that the proximity to the structure make the 40m antenna appear electrically longer than it is physically. Only one QSO has been completed so far and that was with a station in WA. RST report was 469 but WA is off the "end" of the major axis of this antenna. At this time (7/10/97) no contacts on 10m have been completed but the band still appears dead.

Comments

With only one QSO completed, the jury is still out on this antenna. Will update this page as more as information becomes available.

Case 4: 10m dipole in attic

Antenna type

Full size dipole made with #24 gauge magnet wire and ceramic insulators. Oriented approximately NNW by SSE. Actual elevation above ground was approximately 15 to 18 feet. Dipole was not perfectly horizontal, instead it was oriented more like an inverted V with about a 15 to 20 degree slope.

Structure

Single story, single family house. Wood frame, stucco exterior walls with a concrete tile roof. Shack was located in Southern California.

Rig Details

Icom IC725 running average of 50 watts. Coupled to antenna through RG58/U coax. Used MFJ-941E tuner to match rig to antenna.

Performance Summary

No contacts were made. Did manage to hear two or three other stations but all attempts to reach them were unsucessful. Wasn't able to determine how much power the other stations were running or the type of setup used. Also used the antenna as a receive only on 40m, this appears to have worked but without another outdoor antenna that could be switched in, no comparative measurements could be made.

Comments

Data for this case is inconclusive. Band conditions on 10m were very poor so there may not have been any stations there to begin with. Theorized that the concrete tiles could have been interfering with the radiation of energy from the antenna. The antenna has since been relocated outside the house under the eaves - see case 3.

Links to other ham related sites

FCC Home page

FCC Downloadable forms Includes form 610

ARRL Home page

Near realtime MUF map actual color map updated every 30 minutes

AC6V Ham Radio DX Reference Guide Tons of great links

KB9JJA Ham Radio Links another source of great links

WE6W Home page St Louis vertical information

W6MMA upgrades to the St Louis vertical

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KE4UYP'S

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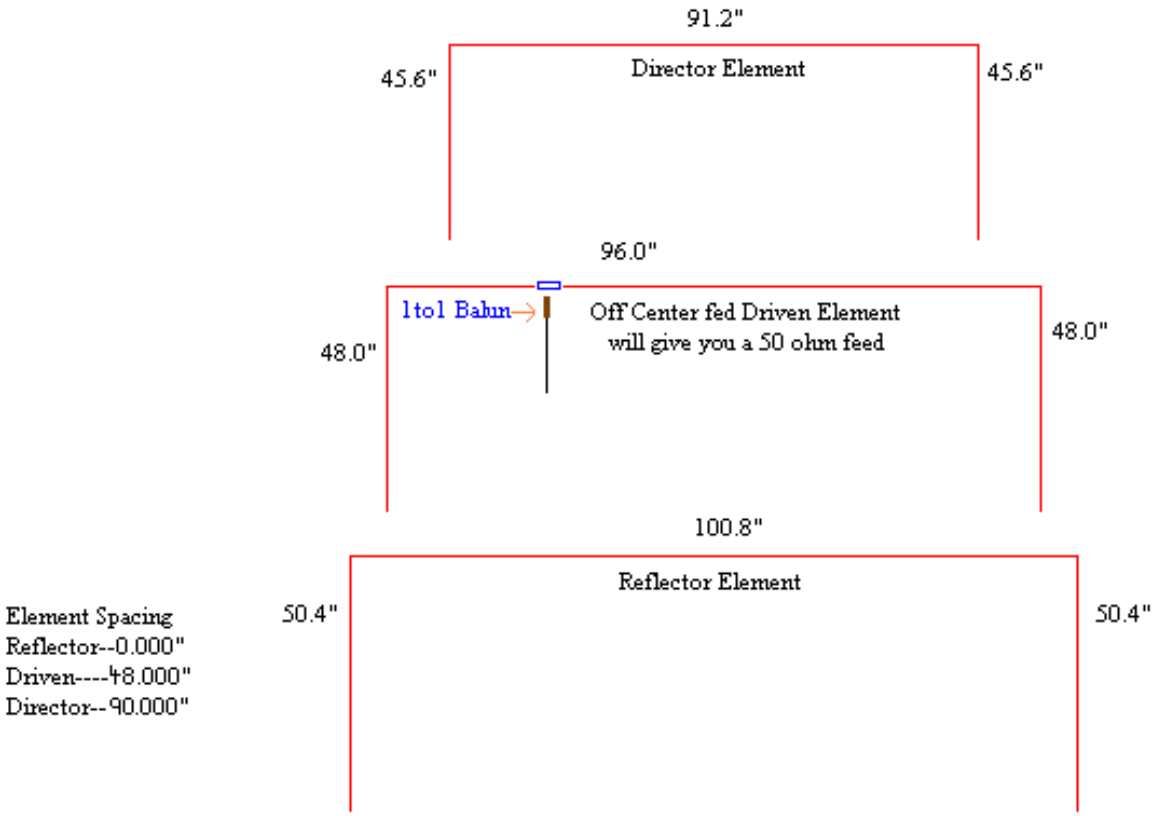

Specifications

This is a horzontally polarized three Element Yagi. The bent Elements cause very little performance lose.
When you consider it is only 8ft. 5" wide and 7ft. 6" long and still has 6.2dbi gain and 20db front to back ratio.
And when you off center feed it you get a perfect 50 ohm match to coaxial cables.The SWR is less than 2to1 for 600khz.
Not bad for a Antenna that the neighbors don't even know exist.

The question
HOW MANY INCHES FROM THE CENTER OF THE DRIVEN ELEMENT IS THE FEED POINT?
Is not a simple one, each installation is different there is no set distance from the center. There is too many variables to consider the height above ground is one of them the gauge wire is another all objects that are within 16 feet is one more but there is a simple solution to this problem. If you make each end of the driven element adjustable where you can easily change the length then you can easily off center the feedpoint to the exact spot that gives you the 50 ohm match.
Remember that you want to keep the total length of the driven element 16 feet long so if you shorten one side by one inch you want to lengthen the other side by one inch this will maintain the 16ft. if you make the last 12" of each end of the element where you have a second wire that is wrapped around the element wire and it is also 12" inches long you will be able to lengthen or shorten that wire easily.
So one-half of the driven element would be seven feet long and the other half of the driven element would be 8 feet long and you would be able to lengthen the seven foot element up to 12" inches longer and you would be able to lengthen the eight foot element up to 12 inches longer once you get your perfect match all you have to do then is solder the sliding wire permanently to the element on each end.

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W3FF Antennas is a manufacturer of a lightweight, versatile hf portable dipole antenna system for amateur radio. This hf antenna is instantly configurable as a vertical dipole or any one of several other shapes including a vee dipole, sloper, or L. The Buddipole dipole antenna is directional and can be converted to a an effective single element vertical antenna (the “Buddistick”) by using either one of the dipole antenna arms and a counterpoise. The Buddipole or Buddistick are perfect for space-restricted covenants such as use as an apartment antenna, hotel antenna or backup marine antenna, or as a compact dxpedition antenna or backpack antenna.

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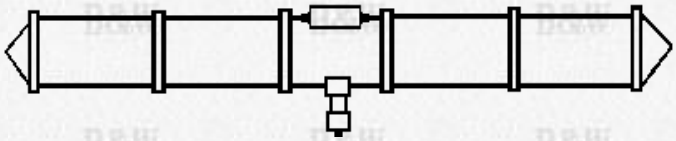

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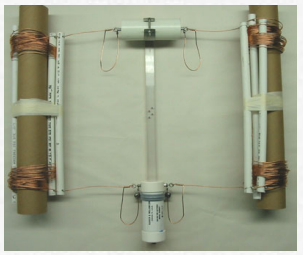
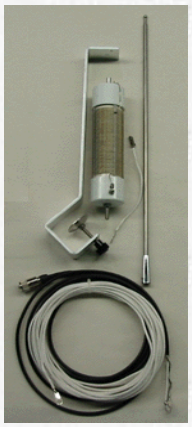


Broadband Antennas - Our field proven broadband HF antennas operate *all* frequencies continuously from 1.8 to 30 Mhz *without an antenna tuner*, plus 6 meters. If you are participating in various programs such as MARS, ARES, RACES or CAP these antennas will work all of your frequencies all the time. If your frequencies get reassigned, it does not matter.

Of course, these antennas are not limited to emergency operation. Any radio amateur enthusiast can take advantage of full coverage with one antenna that has a fairly small visual impact. We even have a configuration that can have the ends run down to the ground - *pretty stealthy!*



Other Antennas - Our AP-10A

window/portable antenna lets almost anybody get on the air regardless of covenant restrictions, and is truly portable. For travel it easily fits in a suitcase or briefcase.

Components and Accessories -

Professional and amateur users alike can make use of various products we carry such as our world famous air wound coils, chokes, insulators, wire, cables - just look around the web site for all of your needs.

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Antenna restrictions? - Short on Space?

The CliffDweller II Antenna

The CliffDweller II antenna!

The CliffDweller II is the Original variable length HF antenna that adjusts it's size to fit YOUR operating environment. 6 to 80 Mtrs. - QRP to 100 Watts -

Although composed of a full 130 feet of wire (a full half wavelength on 80 mtrs!), the CliffDweller II will operate from a compressed length of only 15 feet to a fully extended length of 50 feet -- or anywhere in between. It fits the space you have available!

Fully compressed the CliffDweller II takes up less than one foot of space. Take it with you anywhere an antenna may be needed! Permanent or portable operation...

CliffDweller II - $39.95 plus $10.95 S&H (U.S. only) - *$2.00 Insurance.
*Not responsible for lost or damaged packages - Please insure your purchase.
(Foreign orders please write for International shipping rates)

See main USA 2Way page for ordering information.

***30 Day Satisfaction Return requires antenna to be in resaleable condition.**

*When using this (or any antenna) indoors for transmitting purposes please use low power and keep the radiating elements as far as possible from humans and animals. RF energy can be harmful.

Back to USA 2Way Home Page

~ ~ ~ The original CliffDweller ~ ~ ~

For best multi-band results an antenna tuner is recommended - Items may vary slightly in appearance or construction.

The USA 2WAY

30 DAY Return Policy

The BEST Antenna Guarantee in the industry!

The CliffDweller II comes our ***exclusive*** *30 Day SATISFACTION Guarantee.
If it doesn't work in your environment - send it back! We'll issue a refund **(minus S&H)** in your original method of payment.

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Amateur Wire Antennas

Par Electronics
End-Fedz
Wire Antennas

back

The **Par End-Fedz** are a full length half wave dipoles, but with an important difference. The coax connector is at one end of the dipole, where it is most needed. These antennas can be mounted horizontally, vertically or as a sloper. No ground plane or counterpoise is needed. Portable operation could not be easier. Simply hang the far end from a tree limb -- the coax is at the bottom. Hang it up in a hotel window or string it up in the attic. End insulators are supplied making suspension easy.

The UV resistant ABS plastic housing encloses an efficient matching network allowing the antenna to be fed with common 50 ohm coaxial cable. All hardware is stainless and the SO-239 connector is silver/teflon. The radiator material is a tough polyethylene jacketed, highly flexible #14 copper wire (41 strands of #30 gauge). One end comes with a #10 solder lug making attachment to the matchbox simple and allowing the radiator portion to be replaced if ever necessary. Power rating is a conservative 200 watts (except EF-20/40). Lightweight (the 20 meter version is only 0.5 pounds), they are ideal for portable work. The all black construction makes them difficult to see. Click here to see an alternate view.

Model	Band	Length	1.5:1 BW	Order #	Your Price	Order
EF-30	30 Meters	T.B.A.	T.B.A.	#2836	T.B.A.	
EF-20	20 Meters	33.0 Feet	300 kHz	#4456	$41.95	Order
EF-17	17 Meters	28.0 Feet	350 kHz	#4457	$41.95	Order
EF-15	15 Meters	22.0 Feet	400 kHz	#4458	$41.95	Order
EF-12	12 Meters	17.5 Feet	500 kHz	#4459	$40.95	Order
EF-10	10 Meters	16.5 Feet	600 kHz	#4460	$39.95	Order
EF- 6	6 Meters	9.2 Feet	1200 kHz	#4461	$38.95	Order
EF-20/40	20/40 Meters	39.0 Feet	Information	#2831	$57.95	Order

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  Suite #101
  Code WP
  Fredericksburg, VA
  22401
- VA residents please add 4.5% sales tax.
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E Z HANG

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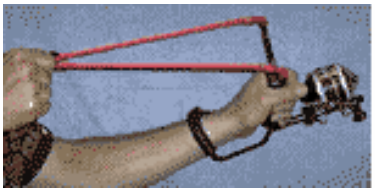
THE *EZ HANG™ WAY*

US Patent No. 6,286,495



- **No climbing trees or ladders**
- **No dangerous bow-and-arrow**
- **No tangled string**

The perfect tool when you want to put up a wire antenna easily, quickly and safely.

EZ HANG ™ can be used to "shoot" a line over a tree or other object over 100 feet high.

Tree surgeons and gardeners can use EZ HANG ™ to attach a line to a tree and pull it safely away from a structure in order to cut off branches.

Cable pullers can use EZ HANG ™ to "shoot" a drag-line through a dropped-ceiling/plenum up to 60 feet and avoid using long pull-rods. It cuts installation time and your costs dramatically. Click Here for the Cable Pulling Page.

Remember! SAFETY FIRST! The EZ HANG ™ propels a high velocity projectile. Always use safety glasses or safety goggles when using the EZ HANG ™. Be sure the area is cleared of personnel before using.

EZ HANG's ™ basic construction is steel attached to a reel that is corrosion-resistant plastic and stainless steel. The reel comes with 300 feet of 10-pound-test monofilament line installed, a quick-disconnect clip to release the weight and an easy-to-see "bright yellow" one-ounce lead weight.

Don't be fooled by "rip-off" imitations that use inferior materials and a poor design!

Look for the EZ HANG™ name.

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Wide Band 155-174 MHz Vertical Antenna (see photo below)

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Force 12 performance, look at 80 /75 mtr stations worked from N6BT with Sigma-280 Ultra at 50' (3/16/04 - including comments about QSO's with YA8G and EA8BH)

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Check out WA5VGI, now 104 cfmd on 160! (6/8/04)

Amateur

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We brought you out of the dark ages of traps - stay tuned NEW! Tornado Tuning Drive

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The Falcon - (project "J-6") 20-6 mtr 6 band rotary dipole

30' FT30-M above in 6' sections. No tools.

15' FT15-T tripod in 4' sections. Goes up in minutes. No tools.

Sigma-164 wide band 155-174 MHz vertical dipole antenna , fully assembled, strong, 99.9% efficient, handles 300 watts power.

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XR-5: the hot 5-bander. In 1 year, Alan, N5PA worked 276 countries!

C-31XR & the Mag 240N in 4 months, Pete, W1RM achieved WAZ & DXCC on 40, 20 & 15 running only 200 watts (99.9% CW).

C-3SS named "Buster" by Alan, K6IPM: 268 countries worked & 240 confirmed in 1 year (April 2002 to 2003).

Team Vertical at Casita Villa located on Discovery Bay, Jamaica

Team Vertical in the 2003 CQWW CW claims 3 more world records: 160 QRP (6Y0A, K2KW op), 80 LP (6Y8A, N6BT op), 15 mtr LP (6Y9A, WA6O op). This will give Team Vertical all 6 QRP world records and 2 LP world records all current. QTH will soon be available for others to use. Read K2KW's write-up about his Top Band work: K2KW/6Y0A

Contact Force 12, Inc. for information on:

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You are free to print all the information in this web site for your own use. ***(Updated 08/03/04 - Happy Birthday Traci - Dad.)***

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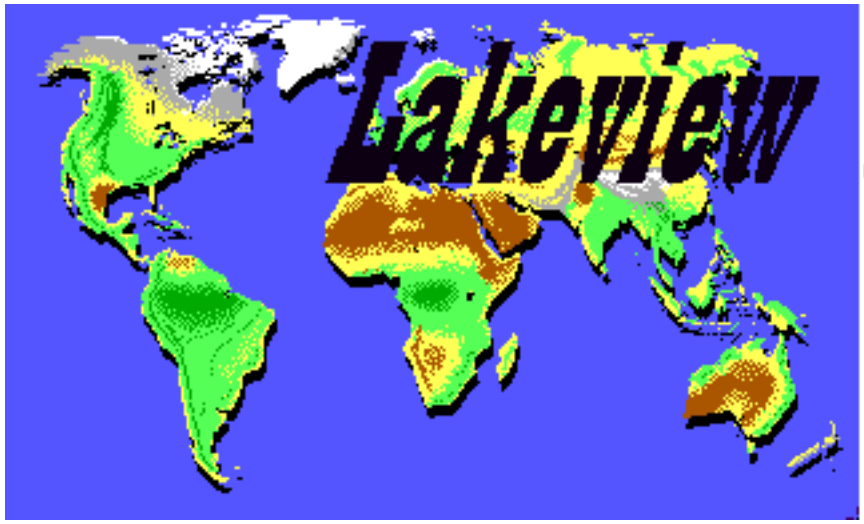

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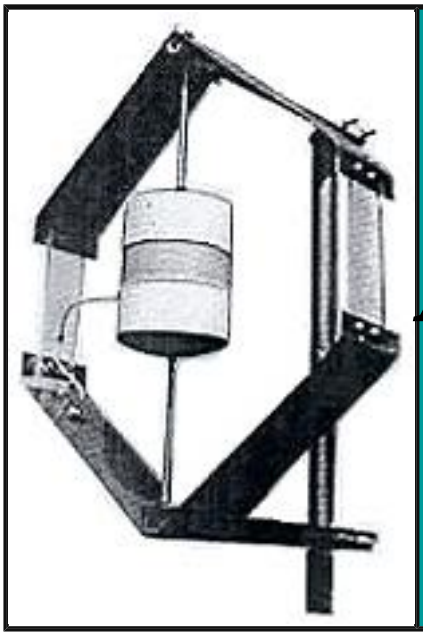

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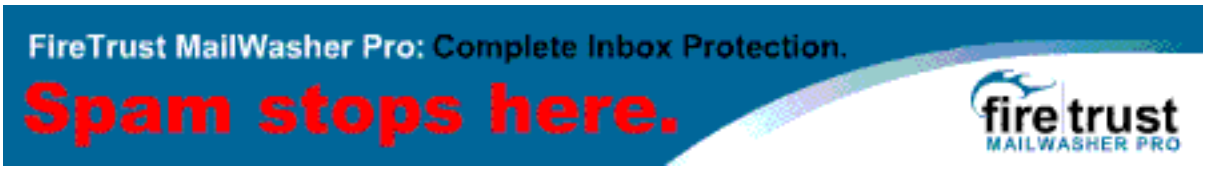

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The MFJ-1622 Apartment Antenna covers 40 through 2 Meters, mounts outdoor to windows, balconies, railings, and works great indoors mounted to desks, tables, and bookshelves!

This Apartment Antenna lets you operate 40 through 10 Meters on HF and 6 and 2 Meters on VHF with a single antenna! Its universal mount/clamp lets you easily attach it to window frames, balconies, and railings. It also works great indoors mounted to a desk, table, or bookshelf. It's not a five element yagi, but you'll work your share of exciting DX!

The highly-efficient air wound 'bug catcher' loading coil and telescoping 5 1/2 foot radiator lets you really get out! The radiator collapses to 2 1/2 feet for easy storage and

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It includes coax RF choke balun, coax feed line, counterpoise wire and safety rope. The operating frequency is adjusted by moving the 'wander lead' on the coil and adjusting the counterpoise for the best SWR.

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MFJ-1952	4.5 Foot Extra-Long Telescopic Whip	14.95
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MFJ-11C06	6 Pairs of Anderson PowerPole Connectors	5.49
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The MinuteMan 20 is, we believe, the best performing portable antenna of its type on the market today. It operates on 20-17-15-12-10 Meters without a tuner (in most cases). On 15, 12, and 10 Meters it's a full 1/4-wavelength long. For 20 and 17 Meters, a small bit of High Efficiency loading coil is added. For more information on why efficiency is so important, and why our antenna is better, go to those pesky laws of physics

It fits in any standard briefcase, backpack or airplane carry-on. We designed it to be rugged, lightweight (about 5 lbs.) and field serviceable. It's perfect any time you need a quick set-up antenna.

Why it's better:

- Sets up in 3 minutes or less, with no tools and no supports needed
- Outperforms competitive antennas
- Fits in a briefcase, backpack, carry-on bag, etc.
- Less expensive, and no "extras" needed to purchase

Specifications:

Height: (when fully extended) appx 12'10"
Power: 100 Watts
Weight: appx 5 lbs
Radials: 4 radials included, each 16'8" long
Longest piece: (when disassembled) 17"
Base Size: (when assembled, base is included) 3' x 3'

Electrical Description:

The MinuteMan 20 is a classic 1/4 Wave vertical groundplane antenna on 10-12-15 Meters. A lower element of #14 wire attaches to a 6' telescoping whip. The whip length is adjusted for best SWR match. For 17 and 20 Meters, a small loading coil is inserted between the elements. The coil is oversquare -- that is, larger in diameter than in length. This Hi-Q design provides much lower loss than the long skinny coils that others use.

Four counterpoise radials are included. Each consists of two conductors, one 8'4", the other 16'8". These lengths are approximately 1/4 wavelength on 10 and 20 Meters. This arrangement strikes an ideal balance between portability and ease of use on one hand, and high earth loss on the other.

To download a printable PDF file of the manual, click here.

Pricing

We sell the MinuteMan 20™ for $130, plus $12 shipping and handling. We now accept PayPal.

You may find it more convenient to purchase through one of our authorized dealers. Each dealer sets his or her own retail price and shipping charge.

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Hi-Q-Antennas™ by W6HIQ
VE7BOC / HA5CMG

Special Announcement!

Visalia, Dayton and SeaSide Ham Conventions are over and thanks to those customers who took time out to see us and to buy a Hi-Q-Antenna or to order one for later delivery. Our back log is more manageable than last year since we have ramped up our production for the Ham versions and have separated the Military types in production.

We INTENTED to EAT the competitors' LUNCH and NOW we will eat their DINNER as well! Not just by the QUALITY and PERFORMANCE, but by PRICE as well! Please visit our website (and theirs) and COMPARE! ASK them for the RADIATION EFFICIENCY and other technical details that pertain to an HF Mobile antenna, and see what kind of answers you get. Ask if they have Hewlett Packard antenna test equipment to verify radiation Field Strength... ask about the COIL "Q"... and ask if they have US or Foreign PATENTS. As our motto says: Shop around and compare, BUT hold on your wallet until you get to the Hi-Q-Antenna products.

My very 73 to you all,
Charlie Gyenes, W6HIQ (HA5CMG)

Welcome to the Hi-Q-Antennas™ Home Page!

The Hi-Q-Antennas Mission Statement:

Force the competitors to produce a better quality and better performing antenna by LEADING the WAY!

The new, improved, Hi-Q-Antennas™ by W6HIQ / VE7BOC / HA5CMG are the latest, state-of-the-art, 6th generation, stealth design. Hi-Q, Hi-Radiation efficiency is provided by computer design and Mil-Spec materials used in manufacturing.

Remember, the Hi-Q-Antennas design is NOT a cheap copy of the original Don Johnson screwdriver antenna. These Hi-Q-Antennas™ products are the **Highest Radiation Efficient** and **Best Quality** HF Mobile antenna EVER designed and manufactured, and are covered by one or more of the following patents: US 6,275.195 B1 and US 6,496.154 B2. For a more in-depth comparison between the Hi-Q-Antennas design and screwdriver-type designs, click here.

Please use the links to the left to learn more about my **ANTENNAS** and **ACCESSORIES**. Also, be sure to check the **LATEST NEWS** page for new product announcements, hamfest schedules, etc. You may also wish to visit our **MESSAGE BOARD** to ask questions and get advice from other owners of Hi-Q-Antennas products. When you've decided on the right antenna for you, utilize the Secure **SHOPPING CART** to place your order.

If you have any questions, please don't hesitate to contact me. You may also get answers to many common questions by reading the **Hi-Q-Antennas™ INSTALLATION MANUAL** and by visiting the ARRL Automotive RFI web site.

73 - Charlie Gyenes
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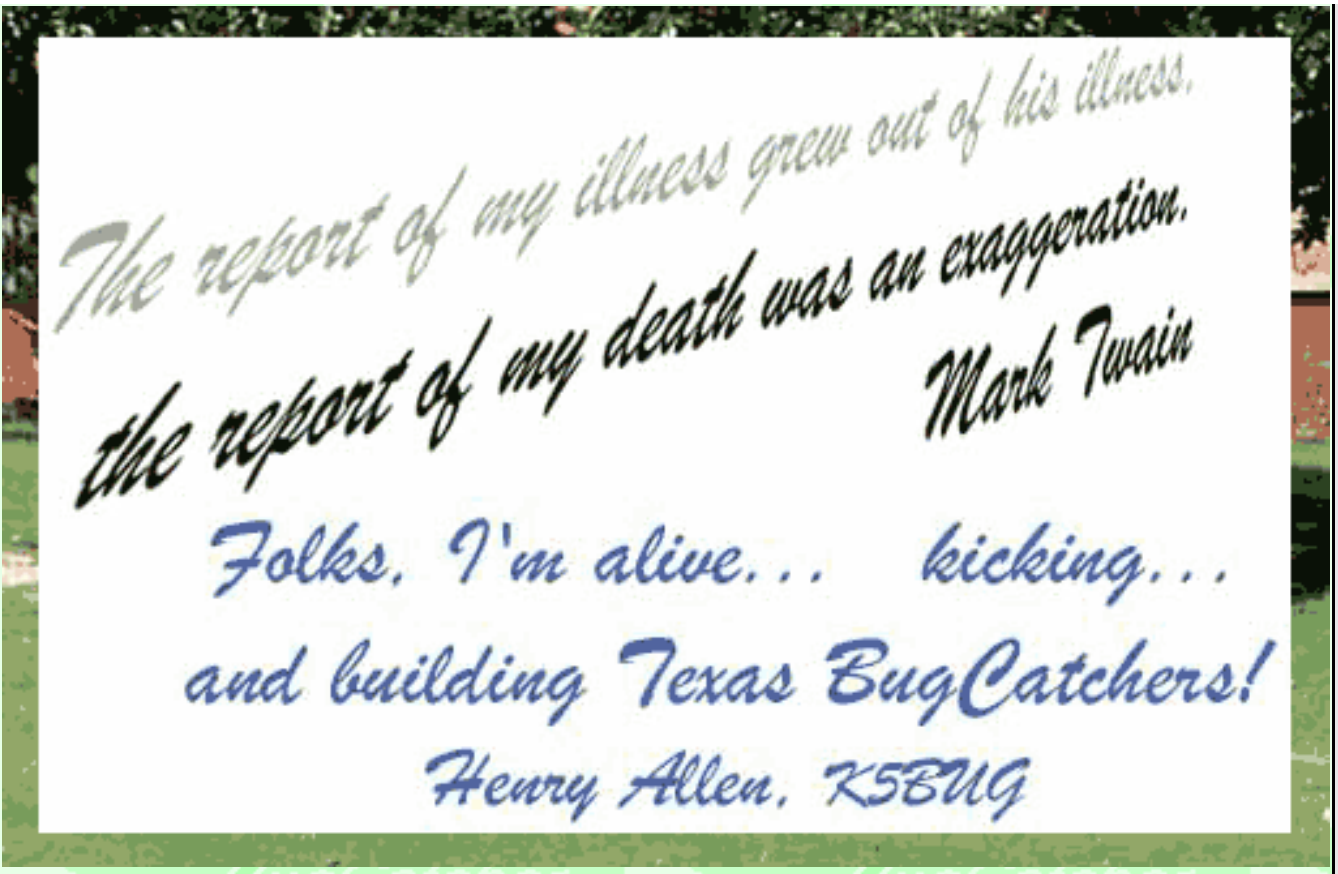

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Amateur Radio Ventennas

Specifications

Model	Frequency (MHz)	Gain (dBi)	Length	Feedline	Price
VT-14	144 - 148	2.6	36 in.	3 ft. RG-58	$54.95
VT-22	222 - 225	2.6	19 in.	3 ft. RG-58	$54.95
VT-44	440 - 450	2.6	16 in.	3 ft. RG-58	$54.95
VT-27	144-148 & 440-450	2.6	36 in.	3 ft. RG-58	$84.95
VT-90	900 - 928	4.2	19 in.	3 ft. RG-58	$84.95
VT-12	1270 - 1290	4.2	19 in.	3 ft. RG-58	$84.95
VT-15S	140-172 & 440-512	Receive only	36 in.	3 ft. RG-58	$54.95
VT-SWL	2 - 30	Receive only	36 in.	3 ft. RG-58	$129.95

Note: all feedlines terminated with an SO-239 connector.

The Ventenna is constructed from rugged ABS plastic tubing. As such, it is virtually indesctrucible. It is designed to fit over vent pipe sizes from 1 to 3 inches, and works over metal or plastic vent pipes. The Ventenna is hollow, thus it does not disrupt the vent function in any way.

The normal "ABS Black" color of the outside of the Ventenna may be painted with any non-metallic-based paint. This allows you to match the Ventenna to the other fixtures on the rooftop.

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Click on the links below to learn more about the *HFp*.

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Specifications and performance are the same as the GIA-XXX series antennas. With its great performance and effective disguise, this is the best clandestine antenna available.

Just in case the neighbors want a Weather Vane like yours, there is a special model just for them. It looks exactly like the STEALTHY antenna but is only a weather vane.

Model	Band	Elements	Price	S & H
WVA-146-2	144-148 MHz	1	$59.95	$5.00
WVA-223-2	222-225 MHz	1	$59.95	$5.00
WVA-445-4	440-450 MHz	2	$69.95	$5.00
WVA-NV	Neighbor Version		$29.95	$5.00

Prices are in U.S. Dollars. Foreign orders add $10.00 S & H.

Don Stoner, W6TNS, has an interesting dissertation on the need for hidden antennas on his web site.

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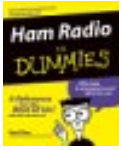

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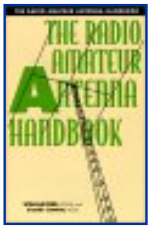

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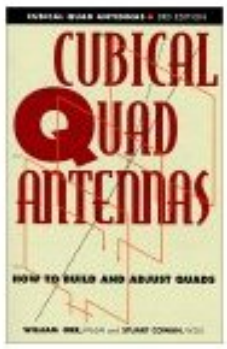

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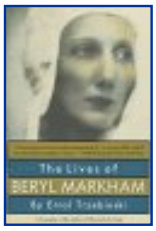

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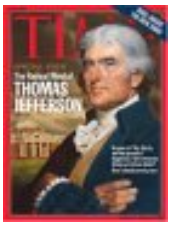

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This book is highly recommended as a basic reference and a good stepping stone to the next level.

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★★★★★ **Simple low cost wire antennas for radio amateurs**, April 10, 2002

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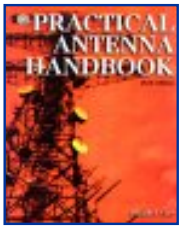

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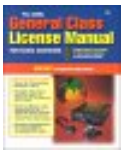

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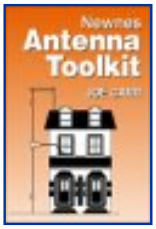

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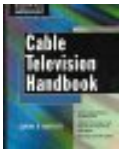

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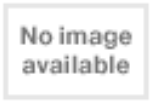

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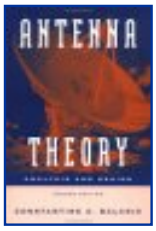

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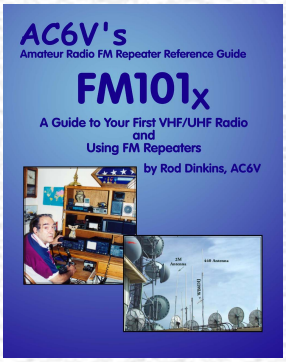

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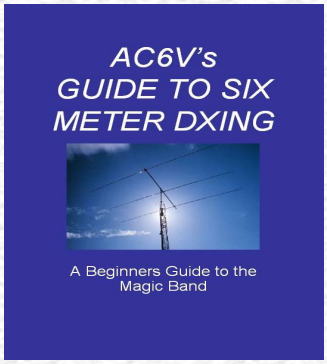

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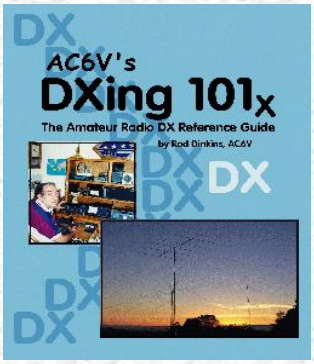

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- ---US CALLSIGNS
  ---73, Q-SIGNALS
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- ---RST Sys & S-Meter
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Hark! I Have Hurled My Words To The Far Reaches Of The Earth! What King Of Old Could Do Thus? --- © AC6V
One May Know Of The Whole World Without Leaving The Shelter Of Their Own Home -- Lao-Tse - 500 BC

73 THANKS FOR STOPPING BY
Counter Stats Thru December 2003 -- 5,719,404 Since 1997 -- Our 7th Year On The Web
Average 3,078 visitors per day
From Over 127 Countries

This Page Last Updated: August 23, 2004 All Links Checked - December 7, 2003

AC6V E-Mail Address

CLICK HERE **CLICK HERE**

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GOTO THE AC6V MAIN PAGE

Check Here If You Want To Open Links In A New Window

THE AC6V BRAG TAPE

AC6V

Rod Dinkins
4982 Marin Drive
Oceanside, CA 92056-4973 USA, San Diego County
Click Here For Map Use Zoom Out For Nearby Cities
E-Mail Address
First Ham Ticket - April 2, 1977
10/10 Number 18029
Grid Square DM13ie, CQ Zone 3, ITU Zone 6
I QSL 100% -- Send Via eQSL Or Snail Mail Or ARRL Bureau

Hi I'm Rod Dinkins living in Oceanside, CA. I started in radio in the late 1940's as an SWL, AM, and Ham band Listener with a Knight Kit OceanHopper - later a Hallicrafters S-40A.

Electronics seemed to be my calling and I had 2 years of vocational electricity and electronics in High School. Four years in the US Navy during the Korean War as an Aviation Electronics Technician - AT1. Also taught electronics at NATTC Memphis, TN - one year.

Four years at Convair Pomona as an instructor of electronics and guided missles. 2 years as a vocational electronics teacher at the Junior college level in Walnut CA. Achieved an AA degree in Electonics under the GI Bill. Over 30 years as an Electronics Technical Writer with aerospace and Hewlett Packard. Licensed in Amateur Radio since 1977.

Today I am Full Time DXing (retired), writing Ham Books and enjoy working on my web pages and being as good an Elmer as I can.

Field Day 2002. 11 Year Old Bryce Kozlowski contacts Australia!
With Assist From AC6V and KG6HBF
Photo By W6VR

AWARDS GIVEN TO THE AC6V SITE

Click On Award To Go To The Award Site

CLICK HERE FOR DA SHACK AND DA BIRD

Some Digital Photos with my Olympus D600L

WHY ANOTHER HAM RADIO PAGE???

Just My Way Of Giving Something Back To Ham Radio

73 Rod in San Diego

The Harmonics

Son Jeff is Programming at Sun Microsystems

Son Steve Is A Video/Media Engineer At Apple Computers

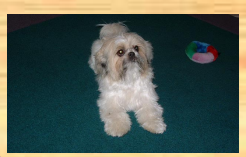
Grandson Ewan (6 years old) -- Has hit the Big Time Music Scene --The Doodle Song -- On line in MP3

Grand-Daughter Rory (3-1/2 Years Old)

The XYL
Karla is an ER Nurse Extraordinaire at the La Jolla VA Hospital

ROD AND KARLA COMMANDEER THE DAWN PRINCESS

K9TZU -- Our New Family Addition



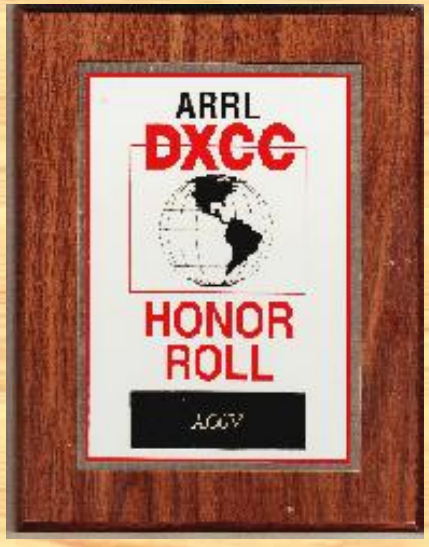
Our Fine Feathered Friend -- VK1BRD

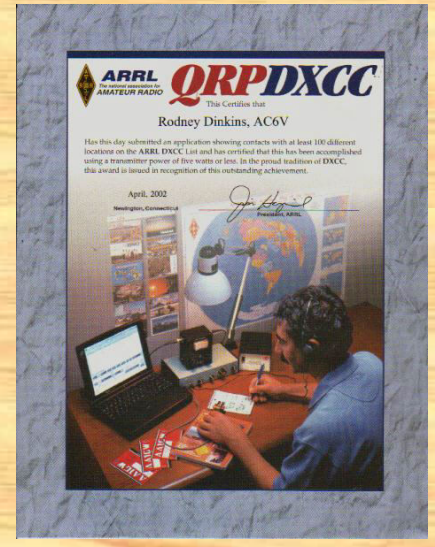
Civil War Re-enactment Photos

AC6V GENEALOGY PAGE

DINKINS GENEALOGY ON FTW

MY HAM AWARDS -- BRAG BRAG





DXCC HONOR ROLL (326 MIXED) ► **DXCC QRP (107 MIXED)**

► **5Band WAS + WAS 17M + 49 STATES ON 12M (Need Utah)**

► **Six Meters -Worked 48 US States and 6 DXCC Countries. 198 Grid Squares (VUCC)**

States Needed For WAS 6M - AR and KY.

► **WAZ 10M, 15M, 20M**

► **DXCC MIXED #18222** ► **DXCC PHONE #8758** ► **DXCC CW #1682**
► **DXCC HONOR ROLL MIXED SEPT 2001** ► **DXCC QRP APR 2002**
► **5BAND WAS #421** ► **WAS CW #27,120** ► **WAS QRP**
► **WAZ 10M SSB #118, 15M, 20M**
► **WPX #888** ► **WAC** ► **WAVE #985**

- **Current ARRL DXCC Entities- 335**
- **DXCC Mixed - All Time Confirmed + Deleted -- 336 (326 Current)**
- **DXCC Phone - All Time Confirmed + Deleted -- 328 (319 Current)**
- **DXCC CW - All Time Confirmed + Deleted -- 168 (165 Current)**
- **DXCC QRP - All Time Confirmed + Deleted -- 107 (107 Current)**
- **DXCC Band Countries - All Time Confirmed + Deleted -- 1085 (1075) Current)**

DXCC is copyright ARRL and its use here is printed with permission of the ARRL.

- **1,745 miles per Watt -- Mozambique C9 Wrked QRP**
- **Ten/Ten Number 18029**
- **Grid Square - DM13ie**

► **WAN -- Worked All Neighbors**
► **WANA -- Worked All Neighbors Again**
► **WAB -- Worked All Beacons (:-)**
► **TORA Award -- Timed Out Repeater Again**

COUNTRIES STILL NEEDED FOR TOP OF THE ROLL (9)

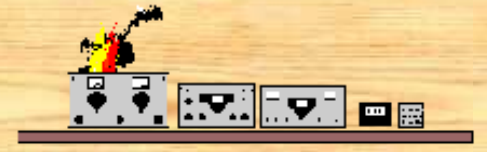
- **3Y - Bouvet -- Not Even Captain Cook Could Find This One!!**
- **7O - Yemen - Missed Em**

- **BS7 - Scarborough Grief**
- **P5 - No. Korea**
- **VK0 - Heard -- I've never even Herd Em**
- **VU - Nicobar**
- **VU - Laccadive -- Lack this one too!!**
- **YA - Afghanistan -- Been Hounding em**
- **ZD9 - Tristan da Cunha**

COUNTRIES NEEDED FOR DXCC HONOR ROLL PHONE (7)

(Or Any Of The ◆ Above

- **9U - Burundi**
- **BV9P - Pratus Isle**
- **E4 - Palenstine**
- **FT5W - Crozet Isle**
- **FT5Z - Amsterdam Isle**
- **VK0 - Macquaire Isle**
- **ZC4 - UK Bases-Cyprus**



But I am QRO!!!!!

U.S. NAVY SERVICE
Rodney Roy Dinkins
USN 4287963

- **US Navy -AT1 - Aviation Electronics Technician 1951-1955.**
  Radio Operator and Avionics Maintenance Technician.
  Air Transport Squadrons - VR-8.
  Aircraft were Douglas R5D and Lockheed R7V-1 Super Constellations.

US NAVY R7V-1 SUPER CONSTELLATION (Photo Credit - William Henry Davis, K1WD)
Aircraft Above Is A Restored Air Force C-121, Navy Version Is Very Similar

Click below to see more photos of the "Connie"

- **The Navy MATS Page -- VR3, VR6, VR7, VR8, VR22 Squadrons**
- **Lockheed Super Constellation -- Navy R7V-1 -- VR8**
- **United States Navy C-130 Squadrons - Including VR-8**
- **Lockheed Super Connie -- The Constellation Group**
- **The MATS Connie Page -- Restoration Group in Arizona**
- **Connie In Antarctica**
- **Blue Angels Super Connie (C-121)**
- **Lockheed Super Constellation Avionics -- APS's, ARC's etc. (Compiled By R. Dinkins)**

- **Air Force Constellation -- EC-121 (551st and 552nd AEW&C WINGS)**
- **Lockheed EC-121 Constellation**
- **The WILLIE VICTOR Roster**

Some Other Navy Aircraft I flew In

Beech SNB -- For Airborne Radar Training (APS-4 B-Scan Radar!)

Douglas Transport Aircraft -- R4D

Douglas R5D - A Navy MATS Photo Another Douglas R5D VR-8

Douglas AQ -- Electronic Counter-Measures Variant of the Able Dog (Spad)

Martin Mars JRM-1 (Mariana's Mars) Flying Boat -- Two are still flying as firebombers

Lots of History and Personal Accounts of the Martin Mars

Photo Gallery USN Planes & Ships I Worked On Or Flew/Sailed In

AMATEUR RADIO CLUBS

San Diego DX Club ARRL Member Palomar ARC

Northern California DX Club - Former DXer Newsletter Editor

AC6V HUMOR

Not For The faint Hearted!!

$ MONEY_MAKIN & JOBS

Sheckle Generation --- Technical Writer - Electronics Instructor

Full Time DXing -- Mar 1999/Present - May Write A Book or Two!

See DX101X FM101x

Solar Turbine -- 1998/1999 - Power Turbines - 18,000 Horse Power!!!!

Palomar Products -- P3V Avionics -- 1997

Cubic -- J-Stars -- 1996

Hewlett-Packard Santa Clara (21 years) -- Freq & Time, Lasers -- 1969-1990

Sylvania -- Reconnaissance Systems -- 1962-1969

Philco -- AF Tracking Stations -- 1960-1962

Convair -- Terrier, Tartar, Mauler, Redeye -- 1955-1960

US Navy -- 1951 to 1955 Didn't make a lot of money there

Republic Steel -- Electrician -- 1949-1951

RIGS & TOYS

HF/VHF/UHF STUFF

Kenwood TS-870SAT Kenwood TS-570S(G) Kenwood SM-230

Ameritron AL-80B Kenwood TS-790A Kenwood TM-261

ADI-220 MHz Icom IC-T2H Kenwood TH-F6A

Bencher Single Paddle Key LogicKey Rascal PSK31

ANTENNAS

Cushcraft R-7000 Diamond U5000 Cushcraft AR6

BOATANCHORS

Hallicrafters S-40B Hallicrafters S-38C Swan 350 (Three-Drifty!)

Kenwood TS-830S Hallicrafters SX-100

SCANNER/SWL

RadioShack Pro 2006 Scanner Radio Shack DX-398 SWL

FRS

Kenwood FRS Radios UBZ-AL14

RAN OUT OF SHECKLES !

Will think of more to brag on Later!!

THE AC6V BRAG PHOTOS

HONOR ROLL -- BRAG BRAG

Latest Shack Shot -- 160 Meters To 1.2 Gig (less 900 Mhz)

Home Made QSL Card Using MS PowerPoint and Printed On Orange Card Stock.
Beam Is NOT Standard Equipment!!

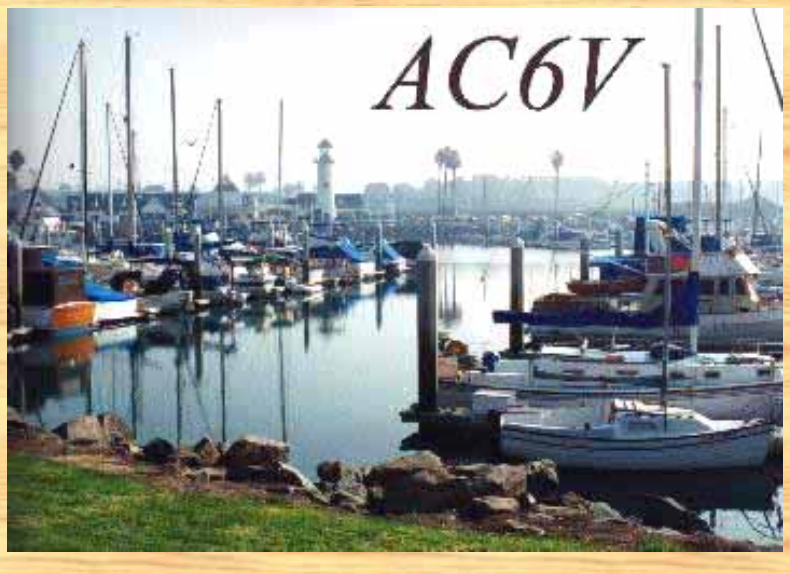


Photo By AC6V With Pentax ZX-5 35mm Camera. At Oceanside Harbor.
Made With an Avery Transparent Label For AC6V Overlay.
Reverse Side Uses an Avery Label For QSL Info.

Star Of India At San Diego Maritime Museum
Photo By AC6V With Pentax ZX-5 35mm Camera.
For callsign - made with an Avery Transparent Label #8660.
Reverse side uses an Avery White Label #5263 For QSL Info.
The world's oldest active iron-hull sailing ship, the Star of India was built in 1863.
Star Of India History 2.

My Latest QSL Card From CheapQSLs

AC6V's Amateur Radio Links **GO BACK TO THE BIG INDEX**

AC6V's Amateur Radio Links **GO BACK TO THE QUICK INDEX**

Last Update:August 04, 2004

Balun and Transformer Core Selection

[Home] [Up]

Related articles at

Balun Test contains model of "perfect" dipole currents.

Sleeve Balun shows how a sleeve adds impedance, useful for VHF and higher baluns

Receiving Common Mode Noise shows how lack of a balun can contribute to system noise (it applies to transmitting antennas as well)

Longwires, Verticals, and Baluns shows how unbalanced antennas can have similar problems

Transmitting baluns on testing transmitting baluns

Occasionally errors are made regarding core selection. This especially includes baluns, where on occasion some very strange ideas surface. One rather odd but somewhat popular idea is that adding a mixture of core types will allow both high power operation and high choking impedance in baluns by slowly reducing current through a balun. Other misleading claims are that extreme values of core ui, such as values in the 10,000 or higher range, are necessary on 1.8 MHz and higher. Other ideas tend support use of excessively low permeability cores for the same application.

When I recommend a core, the material selection is always based on actual measurements with proper test equipment on a bench as well as in the actual end-application.

Core Material

I mainly use 73 material for receiving applications in ***LOW POWER*** applications between .1 and 30 MHz. 73 and similar core materials generally minimize the turns count required without inducing excessive loss. One of the best indicators of correct core selection is looking at the turns required. You'll notice most of the transformers I use have only one or two turns for every 100 ohms of impedance. My 75 to 450 ohm Beverage transformers, for example, only require two-turn primary and 5-turn secondary windings. ***A low "turns count" is a good indicator the correct core size and***

core material is being used.

For high power applications at HF it is often necessary to use lower permeability cores. There are two reasons for this:

- Lower permeability cores generally are available with higher curie temperatures. They operate at high temperatures without losing their magnetic properties.
- Lower permeability cores have higher Q (lower loss tangent) at a given frequency. This means a larger part of the impedance is associated with *lossless* reactance rather *dissipative* resistance. They turn a smaller percentage of power into heat, and that is very important at high power levels.

Permeability changes with frequency. As frequency is increased from zero eventually core impedance peaks. Above the frequency where impedance peaks the impedance of the core (and the ***effective*** permeability) actually decreases.

A downward slope in permeability with increasing frequency is useful for controlling impedance in broadband transformers, but we should be careful to avoid excessive slope. Excessive initial permeability can easily move the operating area too far out on the downward slope of impedance.

A transformer or inductor operating on the downward slope of a high ui core requires extra turns to maintain critical impedance and often requires more turns than a lower ui core. The upper frequency limit will decrease, and this may reduce useful bandwidth in the desired frequency range.

Using excessive initial permeability means winding becomes more tedious (it takes more turns). The wire has to be smaller and more fragile to fit a given core window. Temperature stability is often reduced while losses increase over an optimum core material selection. In addition, stray capacitance increases needlessly, reducing bandwidth and increasing unwanted stray coupling.

Do **NOT** pick cores solely by considering initial mu, since that value is taken at dc. You should consider ***characteristics measured at the operating frequency!***

Always remember this general guideline. Les wire length (as long as winding impedance is sufficient) results in better transformer bandwidth. The best designs place maximum conductor length INSIDE the magnetic core window, and minimum conductor length OUTSIDE the core window.

Heating

At higher power levels, it is necessary to move to lower loss tangent and higher curie

temperature materials like 65, 61, or (in extreme cases) 43 materials. Even a fraction of a dB loss produces significant heating in small cores when power level is in the kilowatt range. *The loss DIFFERENCE in non-resonant applications between lower and high ui ferrite cores isn't significantly different, but* ***heating*** *can be much less!*

We often assume heat means a core is very lossy or is "saturating", but **t**his often isn't true. We must consider the power level, duty cycle, and ability of the core to dissipate heat and look at the full picture.

Very small cores, such as small thin .5 inch diameter cores used on bead-type choke baluns, can only dissipate a fraction of a watt in open air. It sometimes helps to put temperature in perspective by visualizing how hot a 60-watt light bulb runs in normal operation. When we consider the core's size, it usually has significantly less surface area than the bulb. The core also has poor thermal conductivity, and is often stuffed in a container preventing any type of air circulation.

Consider the construction of a typical bead balun, enclosed in PVC and heat shrink tubing. As little as 20 watts dissipated out of 1500 watts can produce damaging heat in tiny beads enclosed in a PVC tube. 20 watts out of 1500 is less than 0.1dB loss, yet it overheats the core!

The problem is almost always a heating problem, and not a core-loss problem. It is almost never core saturation, unless the core is subjected to very low average power and very high peak power levels. It is best that we worry about heat and the number of turns we use, not actual power loss, when selecting a core.

Core Style

Soft-iron cores (soft magnetically) increase inductance because they increase *flux density* near a conductor for a given current. With only a small amount of flux "concentration", there can not be a large increase in inductance or impedance. We need a significant increase in flux to have a significant increase in impedance.

The area outside the core window does NOT have a closed magnetic path surrounding the conductor. ***The presence of the core has a minimal effect on impedance of any conductor area outside the core window.*** Most of the flux from external wires is in air, rather the core. With only a portion of the flux surrounding the outer conductors cutting the outer layers of the core, the ***useful*** impedance contribution of wire outside the core window to system impedance is minimal. Conductor length outside the core window mostly adds unwanted stray reactance and leakage flux. If we MINIMIZE the wire length exposed outside the core, and we generally have a more effective inductor, choke, or transformer.

This effect can be easily conformed in a simple experiment using an antenna analyzer. Connect a short wire across the output of an analyzer, and measure the impedance. Lay a core against the wire,

and observe the very small impedance increase. Now pass the same wire through the core center, and observe the large impedance change. This illustrates why the winding's wire length on the outside of the core is wasted, mostly contributing to undesired effects.

Core Dimensions

The area inside the winding-window of a soft-iron (soft magnetically, not physically) core is cut by all of the flux lines, and this area has a very large effect on impedance. The core concentrates the magnetic flux surrounding a current-carrying conductor into a very small area, and the thickness of the core moving away from the area of the conductor very rapidly has less effect.

- The additional impedance caused by placing a core over a conductor or conductors is almost entirely proportional to the core's internal length (window depth) paralleling and surrounding the conductor or conductors.
- The core diameter or radial thickness only has a small effect on impedance.

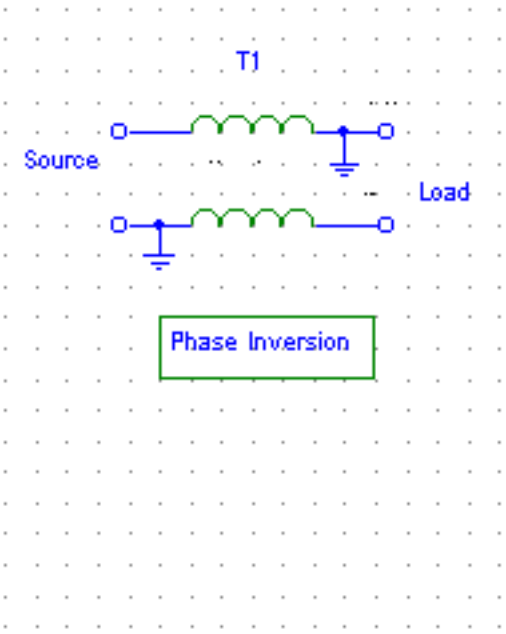
Doubling the core area **parallel** with a conductor roughly doubles winding impedance. The same is NOT true for an increase in core wall thickness, core thickness barely affects impedance.

I prefer binocular cores for most low-power applications and side-by-side stacks of cores (making a large "binocular core") for high power broad-band applications. This type of core arrangement almost always minimizes the amount of conductor hanging "outside the window". With very little conductor hanging "outside" the core window, there is less "needless" wire adding undesired stray capacitance and series resistance. For a given core material and impedance, conductor length can often be reduced to about one-third of a similar impedance choke (or transformer) using a conventional single-hole core or single stack!

Low Power Measurements

Phase Inversion and Choke Baluns

Some of my receiving system designs use phase-inversion transformers. Phase-inversion transformers are identical to (and interchangeable with) choke baluns or line-isolation transformers. For HF receiving applications, 73-material binocular cores are wound with six passes of #26 twisted-pair enameled wire. I use Fair Rite Products 2873000202 cores (about 1/2 inch square and 1/3 inch thick 73 material).

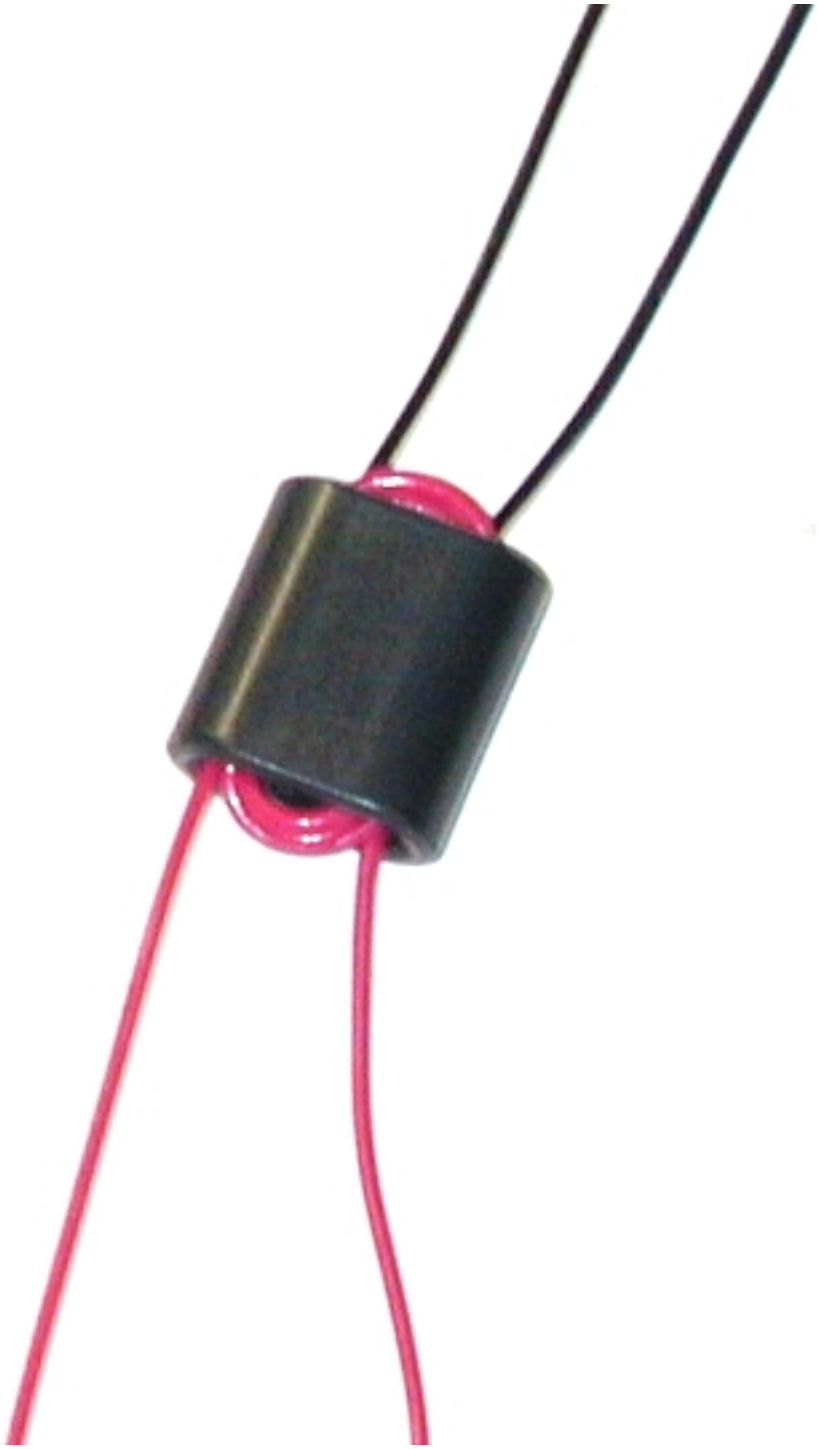


Here are measurements of a sample inversion transformer at 2.5MHz using accurate (and fairly new) commercial equipment:

Load Value (ohms)	Loss (-dB)	Phase Error (degrees deviation from 180)
100	~ 0	<1
33	~ 0	-1.2
10	0.2	-2

This shows inverting transformer construction is good, since even a 10-ohm impedance load works well!

Beverage Matching Transformer



Sometimes I use designs for a long time, and forget how I decided they were OK. I recently received an e-mail questioning the number of turns in my Beverage transformer design, so it seemed like a good time to re-confirm the design.

I retested a 2:5 turn ratio transformer using a single FairRite Products 2873000202 core (about 1/2

inch square and 1/3 inch thick 73 material) two different ways on a generator/ network analyzer/vector impedance test set.

Total loss of two back-to-back transformers was .84dB at 1 MHz increasing, not decreasing, linearly to .98dB at 30MHz. The actual transformer loss would be .42dB at 1MHz increasing to .49dB at 30MHz.

Doubling turns increased the attenuation slope. While 1MHz loss decreased to .69dB per pair, 30MHz loss increased to 1.21dB. This was for a PAIR of transformers connected in series to make a 1:1 transformer. This of course removes mismatch losses, so it is twice the real transformer loss. Actual loss would be .35dB @1MHz increasing to .61dB at 30MHz.

Measuring a second way, I terminated the transformer in 470 ohms. Loss measured .65dB at 1MHz when mismatch loss was included. Since receivers have wide ranges of input impedance, any mismatch error might help OR hurt actual system loss. Factoring out mismatch loss the second measurement indicated about .53dB 1MHz loss with the original 2-to-5 turn transformer and .43dB with twice the turns.

Every measurement has tolerances, and the two different methods do provide different losses because measurement errors affect results differently. Still, it is safe to say doubling turns has a negligible effect on 1MHz loss (which is around .45dB).

Amateur Radio Station K7ZB Arizona

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Jamboree On The Air 2000
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A Stealthy 75 meter Antenna
HF Receiving Loop Antenna
Other activities and photos
Email K7ZB

HF Receiving Loop Antenna

An easy-to-make receiving loop antenna for HF

20m HF Receiving Loop Antenna

Receiving HF signals at my location is a compromise situation due to the antennas I use and the high noise environment in the city. Although my transceiver has good selectivity and excellent filters, I suffer from high noise conditions that at times make receiving a chore. This is true for both weak-signal DX as well as regional chats with friends in the Southwestern US.

I have often wondered whether a loop receiving antenna would be a good solution to my need for better reception but was hesitant to get involved with the typical preamplifiers and monstrous dimensions most of the designs I have seen require. This all changed when I came across a website that described an easy-to-build receiving loop that does not need a preamp. This design does require an external antenna tuner to provide some preselection gain and uses your transceiver's preamp. You simply connect the loop through the feeder to your antenna tuner and feed the tuner into your transceiver's auxillary antenna input connector. You peak the external tuner for maximum signal strength.

Remember: this is a RECEIVE only antenna - it is not intended for transmitting! If your rig doesn't have an internal antenna switchover relay you will need to exercise caution doing it manually. I built my first loop for 40/80m in order to improve reception of signals in the Southwest and inadvertently transmitted once on 75m. I was heard weakly, 150 miles away -pumping 300 watts into the loop - it didn't damage anything but you don't want to operate this way for long!

I built this loop directly from the greertech.com website dimensions (see below) and added my own wooden

support design made from 1/2X3/4 inch stock from Home Depot. I first built the loops and supported them with a simple 1/2 inch PVC pipe mounting system to evaluate them. I liked this antenna so much that the final versions have a simple wood support built into my operating desk for hand rotation. I can remove the entire antenna system in a manner of minutes when I don't want to use the loop.

The photos here show the basic parts and how small the 20m loop really is - each wooden frame piece is 22 inches long. The feeder from the loop to the antenna tuner is 34 inches long, and the loop itself is 5.9 feet in length, per the greertech.com dimensions for RG-58/U coax. If you use a different coax type, make sure you use the appropriate dimensions in the referenced material.

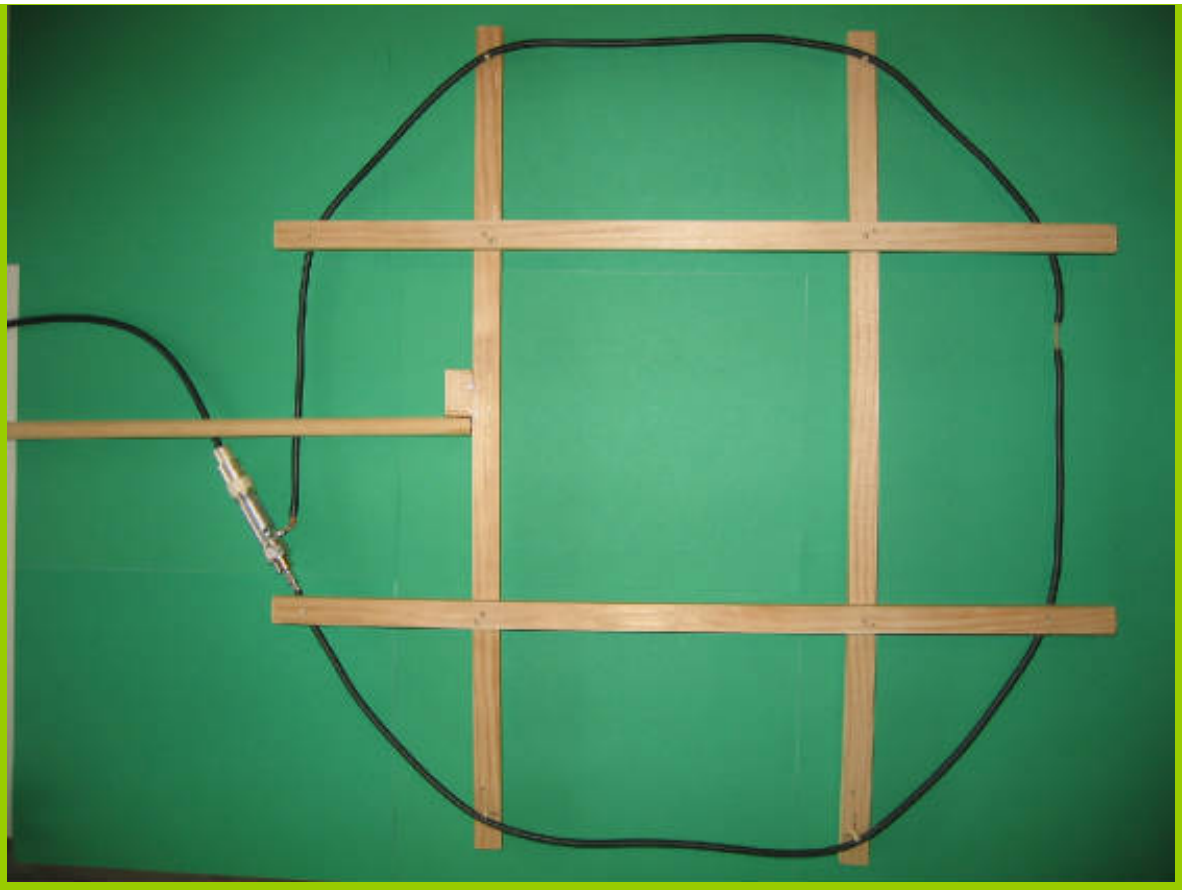
Loop components

I pegged the pieces of the loop support together using dowels so as not to have any metal inside or close to the loop. I made the shorted end of the loop out of an old "blitz bug" lightning arrestor which was in my junk box that conveniently has female '259 connectors and a screwhole in the barrel. Make sure you short only the braid to the center conductor to the braid and don't short out the center conductors! You could simplify things and skip the connector altogether - just solder the center conductor to the braid to the shield as greertech.com indicates on his website. My first 40m loop was built this way and it took about 15 minutes to construct.

Notice the very important gap in the shield braid on the side of the loop - this is one of the key features that makes this design work.

See the greertech.com website referenced below for the electrical details and principles of operation - this is an excellent source of information for these simple, constant-current magnetic loop antennas.

Completed 20m HF receiving loop



RESULTS

This simple and inexpensive loop works remarkably well for something so cheap and simple. It has given me enhanced receive performance in three areas:

1. Nulling noise sources and interference - Orienting the loop so that the plane of the loop is directly aimed at a noise source (and it works for noise sources either near or far, it doesn't matter) allows you to essentially remove the noise. The nulls are sharp, meaning a slight rotation of the loop will pass through it if you're not careful. Of course, if the noise source is in line with the bearing to the signal you want to receive this is not going to help - but in cases where the bearings are different it is a significant help. I am now able to eliminate some sources of noise that have plagued me for years.
2. Peaking desired signals - Aim the plane of the loop to maximize the desired signal and other signals are reduced in strength (unless they are on the same bearing). This loop receives a bi-directional figure-8 pattern, where the peak is in the plane of the loop and the null is broadside to the loop. This little antenna has excellent directivity - I have enjoyed orienting the loop for peak signal strength on both DX as well as US stations and for the majority of cases I can determine the Great Circle Bearing within 5 degrees of the actual bearing. If you enjoy RDF (Radio Direction Finding) pursuits on HF, this is a neat way to do it. There are instances where signals don't show much directivity - I have especially noticed this on Near Vertical Incidence Skywave signals.
3. Lower noise floor - The loop delivers a lower strength signal than a conventional antenna so you need to compensate for this with your transceiver's preamp (or use an external preamp after the antenna coupler and in to the receiver) to bring the signal back up. By doing this in combination with reducing your RF gain and using your AF gain to bring up the signal for comfortable listening you will have reduced the noise floor in your system and will enjoy quieter receive performance. Of course your S-meter won't read like it normally does, but that's a slight penalty to pay when you can hear signals without so much noise!

I am pleased with the improvement this loop has provided in my receive capability. I am able to copy some weak-signal DX I previously could not copy at all and I can now listen on the lower bands with the 40/80m loop and enjoy improved noise conditions.

Here is the greertech website on this loop design:

http://www.greertech.com/hfloop/mymagloop.html

Prototype 40m Loop - works great!

Final 40m version of the Loop...

Antennas

[Home] [Balun and Transformer Core Selection] [Crossfire Phasing] [Detuning Towers]
[Combiner and Splitters] [E-H Antenna] [Omega and Gama Matching] [Radiation Resistance]
[Receiving] [Transmitting] [Transmitting Baluns] [toroid balun winding] [Traps]
[Loading Inductors] [Inductors and Loading Coil Current (Mobile and Loaded Antennas)]
[Mobile Antenna FS Comparisons] [Phasing Systems] [Sleeve Baluns]
[Baluns on Log Periodic Antennas] [Snap on core material data] [Short dipoles and problems]
[End-fed vertical j-pole and horizontal zepp] [Ground resistance measurements]

Multiple Articles at:

Receiving: Contains information in many articles. Receiving antennas including Beverages and Beverage Antenna Construction, Loops, K9AY, Pennant, Flag, EWE, Slinky Beverage Antennas, vertical arrays, magnetic loop antenna, and terminated loop arrays

Transmitting: Contains information on various transmitting antennas

Single articles:

Balun and Core Selection: Contains information on core selection for transformers and baluns

Baluns, Sleeve balun: How they work

Baluns, transmitting: What they do, how they do it, and how to test them

Balun, Toroid: Winding style, debunking split-winding will improve performance

Crossfire Phasing: Contains information on cross fire phasing and why it is superior for broad bandwidth arrays

Combiner and Splitters: Contains information on Magic-T splitters and combiners, how they work and what they do

Detuning Towers: A quick explanation of how it works and the incorrect idea that you adjust for minimum current! Pass this along, it is a major error to tune for minimum current!

EH antenna: the E-H and CFA antennas. How they work.

Inductors, Loading: A brief tutorial on loading inductors

Mobile and Loaded Antennas: Small loaded antenna systems

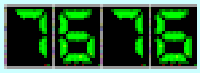
(Related page: Inductor spice model)

Omega and Gamma Matching: Contains technical information on Omega and Gamma matches and matching, impedance limits, component selection, component failures

Radiation and Fields: What the terms we use actually describe

Radiation Resistance: A revised (as of Feb 14, 03) tutorial on radiation resistance and how it is used and misused

Traps: Measurements and operation of traps and trap antennas. Including trap antenna loss.

 **since May 2004**

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[Home] [160 meter History] [Antennas] [Amplifiers] [Amplifier and Radio Relay Sequencing]
[Balun Test] [Checking bandwidth with receiver] [Building a Current Meter]
[Coaxial line and shielded wires] [diversity receiver and transmission] [DX sound files]
[How to Check Signals] [Keyclicks] [MFJ-259B Calibration] [RF Noise Powerstroke Diesel]
[My Shack] [Noise] [Radiation and Fields] [Receivers] [Skin depth and Conductors]
[stacking broadside collinear] [Transmitter Splatter] [Verticals and Baluns] [W4AN]

E-ham

FT1000D Click Mod FT1000D NB mod

W4AN Trust Fund. Bill Fisher contributed a great deal to amateur radio. I hope we can give something back to help his family.

Welcome to W8JI's Home Page

Search:

Google WWW W8JI.COM

All contents of this site are the property of W8JI, but I welcome all links from other sites. Please, if you copy something from here (and even if you rewrite it) add a reference link here! This is entirely to protect technical content, because I refine and update things routinely as errors are called to my attention. Like you, I also learn new things every day. As my knowledge improves, I revise technical articles. I'll note revision dates on articles with changes (if the changes affect technical content).

The most important thing any of us can do to make the web an asset is to help each other with review to insure technical accuracy IMPROVES with time! Let's work to make Internet a reliable source of information instead of a collection of folklore!!

Other Recommended Technical Info Sites:

http://www.eznec.com/index.shtml

http://fermi.la.asu.edu/w9cf/

http://www.seed-solutions.com/gregordy/Amateur%20Radio/W8WWV%20Experimentation.htm

http://www.ifwtech.co.uk/g3sek/

QRN USA Lightning Storm Map

For an instant check on USA mainland QRN click on this link or this link to a weather map of thunderstorms.

Some of this site's (W8JI.com) contents are listed below:

New Page Cebik Dipole and other short dipoles. Feedline and tuner analysis!!

Amplifiers

SB-221 Tank Voltages and the parasitic oscillation myth. The real explanations behind amplifier parasitic oscillations and alleged damage by HF amplifier parasitic oscillations. Look at voltage throughout the tank at frequencies between 1 and 300MHz, and decide for yourself. Related article at Arcs, what makes an amplifier arc?

Preamplifiers for receiving. Measurements of receiving preamps. Compare your favorite pre-amp (text corrections 4/27/03)

Arcs, what makes an HF amplifier arc? This explains causes of amplifier arcing in any amplifier, including the TL-922 amp, Ameritron amps, Heathkit SB220 amplifier, and any other HF

power amplifier for amateur or commercial service. Simple failures and design or operating shortfalls are often explained as parasitic oscillations

Bandswitch failures the real causes and cures of bandswitch arcing and failures in HF power amplifiers that are often attributed to poor parasitic suppression causing oscillation

How to load or tune your HF amplifier and why an amplifier arcs when tuned incorrectly

Inductors deals with antenna loading coils but also applies to amplifiers

Related page: Inductor spice model

Relays how to speed and sequence amplifier T/R relays and how antenna transfer relays affect amplifier failures

Tank Circuits two models showing tank voltages, currents, loss, and phase shift

Vacuum Tubes how they work in RF power amplifiers and why vacuum tubes arc in amplifiers Includes details on 3-500Z and 8877 3CX1500 A7 tubes. Learn reasons tubes fail in amps

Vacuum tubes, how they convert d-c to RF how RF amplifier tubes convert dc to RF power

VHF Parasitics and general stability. Is your amplifier unstable? If you own a TL922, SB220 series, 30L1, or other amps with grid "negative-feedback" you will want to read this! (rev 5PM Feb23/03) Spice models. The resistor think might be non-inductive might not be what you think. Includes internal photos of various resistors.

ANTENNAS and antenna related topics

New Page Cebik Dipole and other short dipoles. Feedline and tuner analysis!!

EH antenna: the E-H and CFA antennas. How they work. See how my early theoretical predictions and descriptions compare to actual field measurements of this most recent antenna theory

hoax! For measurements click here.

If you think ***baluns at the inputs of tuners*** are a good idea, go to this W7EL link and read the section on current baluns. You might be surprised!

Another suggested tuner link is ***W9CF.***

Balun and Core Selection: (mostly receiving) Contains information on core selection for transformers and baluns (it isn't a balum)

Balun, sleeve baluns: How a sleeve balun works and what is important

Balun, transmitting: Testing transmitting baluns and untrue folklore about choke, voltage, and current baluns. Dispelling myth that grabbing coax adequately tests a balun

Baluns: winding methods with toroid cores and debunking myth that a split winding improves performance

Baluns: Common-mode noise. How baluns can help reduce receiving system noise

BALUN TEST: New!!! Test data on transmitting baluns.

Beverages: Construction, Endfire Beverages, Broadside Beverages, Beverage Arrays for low band DX and other low noise receiving antenna arrays for 160 meters and other low frequency bands. Also see balun and core selection for Beverage transformers.

Combiner and Splitters: Contains information on Magic-T splitters and combiners, how they work and what they do

Detuning Towers: A tutorial with an important correction of the common error that we adjust for minimum current!! (revised slightly Feb 17, 2003)

Gain. Stacking (Broadside) and end-to-end (Collinear) gain. How it works, dispelling the 3dB myth.

Loading Inductors: A brief tutorial on loading coils (inductors) and inductor Q

Magnetic Loop Antennas: receiving

Mobile and Loaded Antennas: Loading coils, ground losses, and currents in the system. Related page: Inductor spice model

***Power Stroke Diesel RF Noise*: PowerStroke Diesel noise and how I corrected the problem in my 2003 F-250**

Noise and common mode noise. How it gets into the receiving system Power line and other noise sources

Omega and Gamma Matching: Contains technical information on Omega and Gamma matches, impedance limits, component selection, component failures

Phasing, Crossfire: Contains information on cross-fire phasing and why cross fire phasing is superior for broad bandwidth low-band receiving arrays

Phasing Systems: A quick discussion of phasing methods

Polarization and diversity: Think you can have the best of the two worlds of vertical and horizontal? Think again!

Radiation and Fields: Electric field, magnetic field, Fresnel zone What the terms we use actually describe

Radiation Resistance: A revised (as of Feb 14, 03) tutorial on radiation resistance and how it is used and misused

Receiving and receiving antenna pages: Relative ranking of antennas for Topband DX plus various articles on low-band DX receiving antennas...including low band receiving antennas such as Beverage antenna and Beverage antenna construction. Elongated Loops and other loop receiving antennas such as K9AY, EWE, and flag antennas, small Topband or HF low noise receiving Verticals , my receiving antennas (with a demo), end-fire or cross-fire Beverages, common-mode noise in low noise receiving systems, and more.

Skin depth: The best explanation I have found

Transmitting: Contains information on my various transmitting antennas, including my eight

direction four-square

Traps: Measurements and models of traps, including Coaxial and Tribander Antenna traps

Receivers

Diversity reception "diversity" receiving using stereo along with some sound file examples

How I test receivers

Receiver Tests of some transceivers and receivers. Newly added 756PRO FT1000 MK V Kenwood TS870 (modified)

Receiver IM improvement **mods for FT1000** FT1000D

DX Sound files: What some signals sound like here

1000MP MK V....mods for FT1000MKV receiver IM and transmitter clicks

Key Clicks

Includes sound files of signals with excessive bandwidth, a technical description of what causes clicks, and mods for the FT1000(D) and FT1000MP.

What Causes Clicks....The technical cause and what we can do about clicks.

Keyclick-MP.... An analysis and patch for the FT1000MP 's infamous key clicks

Keyclicks 1000D.... An easy and cheap patch for the FT1000 and FT1000D

Keyclicks 1000MP MK V....mods for FT1000MKV receiver IM and transmitter clicks

Also see Radios.

Noise

Technical article about noise and receiving/ receiving antennas

Power Stroke Diesel RF Noise: PowerStroke Diesel noise and how I corrected the problem in my 2003 F-250

MFJ-1025/1026 Technical Information

Common-Mode Noise issues with feedlines, and how to avoid problems

MFJ-259B Alignment instructions and some technical information MFJ259B.

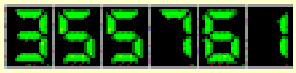
MFJ-1025 and 1026 phasing unit technical information.

Local ARES activities will be included later. We have a local repeater that covers a good bit of I-75 between Macon and Atlanta. Please give a call on 147.225(+600) if you are passing by.

Special thanks to my friend Bill Fisher, W4AN, who is now a silent key. It was Bill's encouragement and generosity that made this site available.

73,

Tom W8JI

This page has been viewed  *times since March 2003. Overall site hits are over 1.5 million for the past year!*

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160 meter History

[Home]

Welcome to a new page that will have archives of 160 meter history. First on this page, a download of W1BB's papers. These papers were contributed by Rolf PY1RO and converted to .pdf files by Ron PY2FUS. Information on this page is intended for private viewing. Publication without permission is prohibited.

History of 160 (about 15mb)

Newsletters

Feb 1962 Bulletin

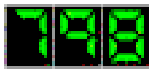
April 1962 Bulletin

October 1962 Bulletin

December 1962 Bulletin

Dec 62 to Feb 63 Letter

1963? Letter (off screen at edges and incomplete)



since May 2004

©w8ji Mar 2004

HISTORY
of
160 METER DX
by
W1BB

This HISTORY is presented to you, by W1BB, in the hope that it will be found useful, and interesting, in connection with your 160 meter activities.

Please advise of any corrections, errors, or omissions noted. Your help in this respect will be very much appreciated.

Best of luck, and HAPPY HAMMING to you always!!

B.C. N. U. on 160 !!!

73!

de

W1BB

HISTORY
of
160 METER DX TESTS
(Dx vs W/Ve)

(The following data for the period from 1931 to 1946 is taken largely from an article by Austin Forsythe, G6FO, Editor, in SHORT WAVE MAGAZINE, for April 1946.)

1931 DECEMBER — G to European Continent contacts. Up to 10 watts power. OK3SK made it with 5 watts power. (3)

1932 JAN-MAR — G to European Continent, Almost as good as 7mc.

DECEMBER — Gs organized tests -- W1DBM with 450 watts, directional 360' antenna on tip of Cape Cod -- Heard s5 to s7 up to 0800 GMT. Peak period 0530-0730.

1933 FEBRUARY — G6FO using 9 watts and 66' antenna with counterpoise, worked W1DBM 0600-0715 for FIRST Transatlantic Crossing. GW5WU also heard in USA

1934 FEBRUARY — Several more Gs worked Across.

1935 JAN-MAR — G6FO-G2PL-G2II-G6YQ, and GW5WU worked across. VE1EA worked G2II and GW5WU first time he ever tried 160 Dx.

1936 JAN-MAR — No scheduled 160 meter Dx tests -- However, FIRST USA/AFRICA 160 meter QSO made by W1BB/FA8BG on 2/23/36. VE1EA worked G2II - G2DQ and G6UJ

1937 JAN-MAR — No scheduled 160 meter Dx tests -- However G2PL/W1BB made FIRST 5-B.T.O.C. within 24 hours, working on 160-80-40-20-10. VE1EA worked across on 160 to G2PL. W2EQS heard in Wales with 30 watts.

1938 FEBRUARY — G6FO - Short Wave Magazine, arranged series of tests. Heard W1BB but could not work across to W/VE - Conditions poor.

1939 JAN - FEB — Band cracked wide open Jan. 8-15-22 and Feb 4 & 16 between 0500-0830. Fifty stations worked across including G-GI-GW-HB-F-FA-SM-OZ-W1-W4-W8. W phones were heard and also identified, but not worked. Interest mounted high. W1BB outstanding W station. Gs across were: G2DQ-2PL-2RC-2CF-2MI-2AO-3GH-3JU-5MP-5RI-5JO-5XH-5QY-6MK-6WY-6QM-6QH-6SQ-6XL-F8ZP-GW2IP-GW8HI-FA8BG.

1940 — WAR --- Band Closed

1941 — WAR --- Band Closed

1942 — WAR --- Band Closed

1943 — WAR --- Band Closed

1944 — WAR --- Band Closed

1945 — WAR --- Band Closed

1946 — Band opened up in "G" land 1800-2000 kc. Local working. No Dx. Loran QRM. Band closed in USA

(The following data from the year 1947 on from the notes of W1BB, other interested Dx Hams , and SWLs.)

1947 — Local G working -- no Dx. Band closed in USA.

1948 — Local G working -- No Dx. Band Closed in USA.

1949 JAN-MAR — Gs heard across again. --- April 9th 1949 W band opened up again by FCC 1800-1825 Kc and 1875-1900 Kc. -- Also other bands for Western USA - See FCC notice.

DECEMBER — VE1EA heard G6BQ and G5JU 12/17/49 11.19PM EST -0415 GMT. G3PU and W4NNN made First Post War crossing 12/18/49 with QSO. GD3UB and VE1EA made second post war crossing 12/19/49

1950 JAN-MAR
-Unorganized tests by leading G and W/VE stations, indicating that band had potentialities for D.x.
-G2PL and G3PU outstanding Dx stations. Other Dx stations working across were: G3FAB-G6BQ-G6GM-G8NF-GD3UB-KV4AA-G2YS. Heard across were GM8UM and EK1AC
-Of special interest - W1BB repeatedly heard very weak signal - almost but not quite identified- called "QRZ QRZ QRZ G6?N G6G?N G6G?N G6G?N - a number of times -- this was later identified as G6ZN with only 3 watts power!!!!!
-VE1EA - W1BB - W4NNN outstanding W/VE stations. Other W/VE stations working across were 10E-3IU-3LII and W2WFZ.
-Of special importance was first G/W two way phone contact on 160 meters between G2PL and W1BB
-W1BB-W4NNN-VE1EA signals heard by FA8BG and EK1AO in Africa.
-FIRST 3 element 160 meter beam tested by W2WFZ with G3PU and with excellent results.
-Eighteen Gs worked EK1AO Jan. 22 -- KV4AA heard by G4LX -- ZB1AR worked several Gs -- HZ1KE worked GM2HIK-GD3UB-G6AB-G6BQ
-W4NNN (Now W9NWX) worked HC1PK during 1950 ARRL DX contest for FIRST USA/South America 160 CW QSO. (Per QST July 1953)
-HZ1KE QSO'd by G. First G-HZ??? First Asiatic??
-TA3FAS QSO'd by G2pl - FIRST G-TA on CW and PHONE.

1951 JAN-MAR
-Organized tests -- by G6FO Short Wave Magazine - and W1BB announcements by QST.
-This was a very successful season, summerized by the following "Outstanding events" from W1BB's report as follows:

#1-First crossing this seasonby VE1EA working G6BQ on Dec. 31, 1950. G6BQ 349 and VE1EA 449
#2-First N.A./Africa Postwar QSO VE1EA/EK1AO Jan 7 - both RST 449
#3-FIRST!!!! Contact between Asia and N.A. on record. between VE1EA and HZ1KE Jan. 14 1951 - 229 both ways.
#4-VE1EA has worked 4 continents - 8 countries Post War - total of 10 countries all time.
#5-W1BB also worked 4 Continents - 8 Countries post war total of 11 Countries all time.
#6-W1BB worked South America HC1JW Jan.21 '51 BB was 469 and HC1JW was 567
#7-W2UKS QSO'd N.A./S.A. same date 349 - W8BKH also. w/HC1JW
#8-W1BB and W9CVQ both heard in NEW ZELAND by ZL1AH Ex G6AH and ZL1MP Ex GW6AA-Ex G2II of early 160 meter DX fame. These reports confirmed. Also an unconfirmed repor of W2QHH being heard by ZL1HM. Howy using 20 watts and 12' high antenna.

#9-A number of Phone/phone and CW/PhoneQSOs were made. G2PL/W1BB p/p F8PJ/W8FNS p/p W1LYV/GW3ZV p/p W1LYV/G2PL cw/p

#10-OUTSTANDING and Most Consistent DX station EK1AO - 300 Watt

#11-Also EXCEPTIONALLY OUTSTANDING for 10 Watts!!!! GW3ZV using miles of wire for an antenna!! Signals up to 589 at times!! This performance deserves a lot of credit.

#12- More US districts worked than ever before!!! W-1-2-3-4-8-9- VE-1-3

-DX working across were: G2PL-YS-MM-ACV-FLK-DA G3PU-SU-DIY ERN G5HB-RI-UB-KM-JU-GU-VB G6GM-QB-BQ-LB G8JM-NF GD3UB EK1AO HZ1KE GW3ZV-FSP HC1JW F8PJ KV4AA.
Heard but not worked were the following stations: G2DPV,NP and G3GGN

-W/VEs working across were: W1BB,EFN,PLO,OMP,ZE,LYV,BEU,SAN, SS,KDX,DVS,RQR,BEU,NRE,AXA,MQB --- W2PTV,ESO,UKS,BQS,BFA,BLR, PEO --- W3LII,FNF,PMG,LOE,EIS --- W4KFC --- W8WKV,FES -- W9FIM,CVQ,PNE, -- VE1EA -- VE3AAZ. Heard also were the following W1AW - W2WC - W2KZX W3OKU,AAA,JAA -- W8FIR,FLH, FLV and WØTQD.

-FIRST!!!!! W9CVQ worked ZL1AH December 31, 1951 RST 229 both ways - after a series of tries and schedules, making FIRST W/ZL 160 meter QSO on record!!!!

-This was an outstanding year -- But??? The SUNSPOT cycle in on the down grade - lets see what happens next year!!

1952 JAN-FEB

_Band was very much poorer than last year -- Due to Sunspot cycle being on downgrade????

-Outstanding events were as follows:

#1-VE1EA made first crossing working G3PU Nov. 25. 1951

#2-W1BB,W9NH,W2QHH and KV4AA heard in New Zeland by ZL1AH and ZL1HM during tests. SPECIAL NOTE!! W2QHH using 6L6, 17 watts, 270'wire 12' highat one end and 3' at the other heard by ZL1HM and CONFIRMED!!!

#3-W1LYV was heard in RUSSIA - received SWL card!!

#4-VE1EA worked HZ1KE for second N.A./AFRICA 160 QSO.

#5-FIRST!! W9/G QSO on record when W9CVQ worked G5JU and GW3FSP December 23rd. 1951 -- but included in 1952 "Season" report here

#6-W2QHH made FIRST VP4 QSO w/VP4LZ on February 23. (Again using only 17 watts.)

#7-W2ESO, Gene Black put weekly reports of test results on Voice Of America broadcasts with special note of FIRST W/ZL QSO

-Dx stations working across were: G2PL,AJ. - G3PU,ERN,COJ,DIY. G5JU - G6GM,BQ- GW3ZV,FSP. - EK1CW - KV4AA - ZL1AH - VP4LZ. ALL Dx heard was worked except GM3OM who was heard and not Wkd.

-W/VEs working across were - W1BB,LYV,SS. - W2TRK,BQS,K2USA - W2RGO - W9CVQ,NH - VE1EA and VE2WW

-It will be noted that while there were several outstanding events, contacts to Europe wase way down, and it was seriously considered giving up the tests for 1953. It was W1BBs suggestio that they be continued to see what would happen.

1953 JAN-FEB

-In spite of poor conditions, there was a lot of activity and interest and good contacts were made. This was due in some measure to better antennas, beter receivers and MORE ACTIVITY.

-The OUTSTANDING EVENTS for 1953, were as follows:---

#1-W1LYV made first crossing for season Nov.9th 1952 working G6BQ, GW3FSP, G3HYG and G3GGN.
#2-FIRST!!! W5/G QSO between W5ENE and G5JU January 4th. G6GM and G3PU were also heard by W5ENE.
#3-FIRST!!! ~~NORWAY~~/USA QSO between OH3NY and W1BB JAN. 11th 0600GMT. W1BB using KYTOON (Balloon) antenna 260' vertical BB/449 NY/339
#4-NEW COUNTRY, EI9J on for the first time - Thanks to W2QHH - with outstanding signal and worked many W/VEs - Had special permit from his government for operation during the 160 meter DX tests.
#5-W1BB,W3EIS heard in South Africa by ZS3K
#6-W1BB,EFK-W2EQS-W3EIS-W9PNE-W9NWX all heard by ZL1AH varying RSTs 229/259 - and various dates during tests.
#7-W1BB experimenting with 260' vertical KYTOON - (Helium Balloon) supported antenna - slightly better than horIZONTAL ANTENNA _ BUT NOT OUTSTANDING _ WEATHER too much of a factor to make it practical. Hope to run further tests for confirmation.
#8-FIRST !!! ASIA/USA QSO on record - W1BB worked ZC4XP on March 8th.
#9-VE1EA worked ZC4XP for his third ASIAN QSO - having previously worked HZ1KE twice.
#10-Several QRP contacts of note were made:
(a) G6ZN with 3 watts worked K2ANR
(b) W2QHH and W1LMU with 17 watts each worked across
#11-A number of phone stations were heard and worked across
#12-First Post War W/EI contact - W2WWP/EI9J Jan 11th.
#13-First Post War VE/EI contact VE1YW/EI9J
#14- W9NH completed FIRST 160 meter "WAS" -(Worked all USA States) - on record. !!!

-Dx working across were: G2HDT,FGD,SU,JF-G3PU,GGN,BKF,FGT,GZK, ATU,GLW,-G5RI,JU,TN-G6GM,CJ,LB-G8KP,JR. Other countries working across were EI9J - GI2ARS-KP4DV-KP4KD-VP4LZ-GW3FSP-OH3NY-KV4AA
--Heard across were G2PL-G3GKQ-G3US-G5VB-G3HVX-G3FEW.
-W/VEs working across were:- W1BB,LYV,LMU,AHX-W2EQS,WWP,QHH,WC. W3EIS - W5ENE - W9PNE - VE1EA -VE1HJ - VE1YW.
-W/VEs heard only were - W1QJM,HSC,AYG,SS,TCR,OJM,FYS,QCA,RQR, DWO,AW,OE - W2KNZ,JPW,HCW,ESO,AMC,NC-W3HL,AVL,TBG - W4VFL,LRN W8NJC,HFG,GDQ - W9NH,MFV,FIM,CZT,BQQ - VE2AIE.
-January 11th was the outstanding dsy of the tests - Dx was gOOD _ QRN was low -- later tests were QRN'd out with terriffic QRN, coinciding with the great hurricane and floods in Britain and Holland.

APR-SEP -SPECIAL NOTICE!!!
-In April W2WWP received word from ZL1WW via ZL2ACV on 40 Cw that 160 meter faint signals were being heard, and asking for schedules.
-On May 3rd W2WWP 1899 Kc QSO'd !!! ZL1WW 1904Kc RSTs WWP/349 WW/239, for SECOND ZL/USA 160 meter QSO on Record!!(See W9CVQ Dec. 1951 for FIRST ZL QSO) -- This was FIRST W2/ZL QSO! - Times were 0900-1030 GMT -- Sunrise this end -- early evening ZL time.
Thereafter W2WWP worked ZL1WW a number of times.
-W2WWP invited W1BB and W9NH to join in on further tries.
-W9NH heard first ZL on July 1st 119rst
-W9NH next worked ZL1WW on July 29th 1022 GMT - MH/449, WW369
-On the same date W2WWP also worked ZL3RB.
-W1BB worked ZL1AH Sunday August 16th at a different time, 0605 for a short "Contact" type QSO. (Sked suggested by ZL1AH)

-W1BB worked ZL1WW, ZL3RB and ZL1AH all on the following Wednesday August 19hh 0945-1030. ZL1WW was a solid QSO. RB and AH short "Contact" report QSOs.
-W2WWP the same morning, as signals were unusually good, almost had a phone QSO with ZL1WW when signals peaked - they heard each others modulation but no phone QSO -- CW QSO again OK..
-W1BB exchanged reports with ZL1NX on August 23rd - also Wkd WW
-Tests are continuing (At the time of this typing) during the summer months and fall of 1953 to see if enough can be learned to establish regular tests in which it would be worth while for all Hams to participate - lots of lost motion this summer.
-Many Many thanks to ZL1WW, ZL2ACV and W2WWP for starting this thing off.
-FIRST !!! 160 meter "WAC" on record made when W1BB QSO'd ZL1AH, ZL1WW, ZL3RB, ZL1NX.

-W2WWP the same morning, as signals were unusually good, almost had a phone QSO with ZL1WW when signals peaked - they heard each others modulation, but no phone QSO resulted. CW qso again OK.
-W1BB exchanged reports with ZL1NX on August 23rd - also worked ZL1WW
-Many Many thanks to ZL1WW, ZL2ACV and W2WWP for starting this ZL wk
-Thus the FIRST!!! 160 meter "WAC"-(Worked all Continents on 160) on record was made when W1BB, QSO'd ZL1AH, ZL1WW, ZL3RB and ZL1NX.

1953 OCT-DEC
1954 JAN-MAR

-This "SEASON", the organized tests were held per G6FO, Short Wave Magazine, and W1BB with announcements in QST/CW and "Break-in".
-There was more activity than ever on 160 - more C ountries, and all 6 Continents heard from.
-Much DX was worked in between the "Organized" tests, and it looks now as though, through these "organized" tests, that the band is being opened up more and more to general operation, and year/mmmm round DX efforts.
-Summary of this seasons OUTSTANDING EVENTS from W1BB's log follow

#1 W3RGQ arranged special tests starting October, and the FIRST CROSSING this season was on Oct 4th when W3RGQ worked G3PU. W1LYV Contacted G3PU later the same evening/ Signals 239/449
#2. FIRST!!! ZL/G QSO!! - 12,000 Miles, October 10th 1953 0640GMT. ZL1AH worked G6GM. AH339, GM339 Half Hour QSO. On the following night, OCT 11th another QSO was made, this time 559 both ways. On Oct 21st G6GM QSO'd ZL3RB - RB439/GM449 -ZL1AH QSO'd G6GM again also.
#3. FIRST W/HB QSO on record -when W1BB worked HB9CM Dec 20 '53 W1BB 449 HB9CM459 0637 GMT.
#4. FIRST Canal Zone/W QSO on record!!! When W3RGQ worked KZ5DE, January 16th 1954
#5. LOWEST POWERED G QSO this season. G2HKU, 5 watts worked W1BB Jan 3rd. 0640GMT - Isle of Sheppy. HKG/329 BB 579
#6. FIRST Hawaiian/New Zelamd QSO. KH6MG/ZL1WW March 24th 1000 GMT
#7. Some FIRST KP4KD QSOs. - EI9J Jan 31st. G6BQ, Jan31st. VP4LZ Feb 21st -VP6EB Jan 31. - VP7NM Jan 31 - VP9BDA, Feb 28th - KZ5DE Feb 7th VE 1EA Feb 14th.

-In general, the first part of the TESTS had better conditions th than the last.
-An unusual phenomonon was observed in that the path seemed to shift from good E -W conditions early, to a more NE -SW path, as evidenced by Canal Zone QSOs with W and G stations

#8. FIRST G/C anal Zone QSO when G3PU worked KZ5DE on Jan 16th

-We had TWO FAKES this season VR2BJ and LU4DM who fooled several.
-DX WORKING W/VE, were:-G3PU, ERN - G5JU, KM, RI - G6GM, QB, BQ, GJ - G2HKU - G8JR - GC3EML - GW3ZV - EI9J-HB9CM-HB9T-KZ5DE-KV4AA-KV4BB-KP4KD-VP4LZ-VP7NM-VP9BDA.
-DX HEARD were G3HVX and G3BKF.
-W/VE WORKING DX were:- W1BB, DBE, LYV, TMA, VDB, EFN, AQE, UJP, ZL -- W2GGL, EQS, AMC, VC, K2ANR, BWR, QHH, WH, PP, UKS - W3RGQ, EIS-W4KFC-W9PNE W0NWX-VE1EA-VE2AIE0VE2WW.
-W/VEs HEARD only by DX were:-- W1AHX, LMU, QCA-W2JIL, MCU, TRK, AMC HEY, KV -, W3FWF, TBG, HL, PA.-W4BRB, POB, KMS-W5ANO-W8GDQ, HQ, IKN, EJL - W9EWC, CZT, NH.
-SWLs!!!!!! - Special Mention should be made of the splendid work done by the SWLs. To list a few of the most consistent ones, --

(next page)

-Norman Smith, BRS20601 -- ROBERT IBALL BSWL 4648 --- George Allen BSWL 4588 -- C.L. Bradbrook G3692 --- William Iball, BRS-20410 -- Mike Birch ISWL/DL2-4832 --- Jack Burgess, ISWL 2561 -- E.M. Hainsworth --- John Whitington ISWL-G3719.BRS 19771 -- Gerald Heslop RSGB A1167/ISWLG4555 --- Ian S Davies RSGB/A1182 --- John L Hall, BSWL/2333. There were others.
-These very factual and accurate reports assisted greatly and added much zest to the 160 meter TESTS. Many many thanks to these cooperative, and fine fellows - the "SWLs" !!!

#3. FIRST!!! - GC/W QSO on record, when GC3EML made contact with W1BB December 17th, 1953. GC 3EML rst 229 - W1BB rst 459.

1954 SEP-DEC
1955 JAN-APR

-This "SEASON" there was a variety of DX, with the North/South path predominating.
-LORAN became more of a problem, with a new station moving in permanently on 1900Kcs, and an intermittant station on 1850. This causes considerable QRM, and gives a high artificial noise level that many weak stations cannot get through.
-Summary, of the OUTSTANDING EVENTS from W1BB's log, follow:-

#1. Season's FIRST USA/Oceania QSO when W8ANO worked ZL3RB Sept. 5th
#2. On Sept 5th G3PU heard W3RGQ 239-349 - and then on Sept. 19th W3RGQ almost worked G3PU for a first season's QSO - when they heard each other briefly through static Crashes, but did not call it a QSO.
#3. First Transatlantic QSO of the 1954/1955 season when W2BQS and G6GM worked each other October 10th 0532GMT. Second QSO of the season - same date - when W1BB worked G6GM 0607gt
#4. G6GM and G6CJ worked ZL3RB again in Mid-October
#5. FIRST VO/W QS O!!! When W3RGQ hooked W4KVM/VO6.
#6. FIRST LU/W QSO!!! When W1BB worked LU3EL November 28th at 0707 GMT. EL/559 BB/569. This was followed by LU QSOs with W8GDQ,W9PNE,WOCVQ,WØNWX,W4ZQ and W3FBX.
#7. FIRST YV/W QSO!!! - Was between YV5DE and W9FIM November 20 1954/ -- Then followed QSOs with W8GDQ,W9PNE,W1BB in order, the next day Nov. 21st. We are greatful to "Mick" for QSOs.
#8. First HB9/KP4 QSO when "Ev" QSO'd HB9CM on 2/20/55 - this makes 17 Countries for "Ev"
#9. FIRST (So far as we know) USA/S.A. Phone QSO, when WØNWX worked YV5FH March 13th.
#10. DXPE DITION!!!! TI9MHB, provided a thrill on 160 and another First, when opening up from COCOS ISLANDS 2/20/51 0500-0800 GMT, the following were worked. A new Country, with WØNWX for a FIRST!!! - He also worked in order -- W4KFC-3EIS-1BB-8DUY-9NH-8ANO-9PNE-3KLA-2QHH-1ZL-TI2BX-9FIM-9CVQ-6KIP/6-YV5DE and 3RGQ.
#11. First YV/g - or YV/Europe QSO February 6th 1955 when YV5DE and G5JU QSO'd.
#12. FIRST!!! USA/Columbia QSO, March 20th when W1BB worked HK4DP at 0540 GMT, followed by W1ZL and W8ANO. Incidentally, this was HK4DPs First EVER QSO on 160 meters.
#13. FIRST!!! USA/HONDOURAS QSO March 27th. W1BB/HR1LW at 1015 GMT through terriffic QRN. Followed by W3RGQ at 0447. This was also "Loui's (Hr1LW) First CW Contact - and also First QSO EVER on 160.
#14. FIRST!!! USA/FAR EAST QSO, when W6KIP/6 worked VS6CQ April 3rd 1955 5.05-5.25 AM PST, at DAWN. VS6CQ using only 10 watts has been trying for USA QSOs 3 years - the only 160 meter signals heard in that time were VS6CZ, a ZL and then W6KIP/6

#15. FIRST "WAS"/PHONE made by W8GDQ - F.B.!!! This is Second 160 WAS. (See No. 14 1953)
#16. SECOND CW "WAS" - and THIRD all time WAS, Made by W2QHH

-W/VEs working DX were:-- W1BB,AHX,VDB,LYV,ZL,ORP,SFE/1,ERX,UXS, YAA,ME,BIL,WY. -W2EQS,GGL,VHH,WVP,K2EDL -W3EIS,RGQ,KLA,DGM,VAN,FBX W4VNE,KFC,ZQ,LXA,INV,LCV - W6KIP/6,VBY -W8GDQ,ANO,KIA,NSF/9,PEV, DUY,PCS,WCL - W9PNE,FIM,CVQ - W0NWX -VE2WK,WW .
-Other W/VEs HEARD only by DX were;-W1BMW,WM/2,CF,LC and VZE -W2MX,P PP,HX,SKE,K2BWR,BVN,WH,WZ,CF,DLO -W3LPN,KMS,RWH/4,HIH - W5CZ - W8 W8PCS,BMJ,SHP/1 - W9CZT,NH - VE2NI - VE3AAZ
-DX working W/VE were:- G2AJ,PL,HX,NX,NN -G3PU,[illegible],GGN,JOJ,JVI,BKF, ERN,JEA -G5JU,RI - G6GM,QB,LB -KP4KD,KP4CC - KV4AA-HB9CM-OK1HI,LM, AJB -OK3AL,VV,DG -OK2AG -ZL3RB,[illegible] -ZL1BY -ZI9J -GI3IOS - TI2BX - TI9MHB- LU3EL - YV5DE,FH-VO3X - W4KVM/VO6 - VP7NM,NG.-HK4DP -HR1LW XE 2OK-KH6IJ
-DX HEARD only by W/VE ---G3HRW, G6QB,LB -OK3VV -YU1GM -TI2WR -

1955 JUN-AUG -SUMMER 160 meter DX!!!!! -Some of the boys believed that they could dispel the myth that DX on 160 was not possible during the summer months. (ZL schedules had been successfully worked in the summer)
-G3JOJ,G3GGN and G3ERN started the ball rolling to test this out, by writing several Ws for skeds during June July August.
*The Ws responding were W1BB,K2BWR,3RGQ,9PNE,9NH.
-On July 3rd W1BB's signals were heard by G3JVI at 579 at Sunrise time G land Sunday Morning. G3JVIs signals were weakly heard.
-The following Tuesday morning K2BWR worked G3GGN 1830KCS for the FIRST!!! summer crossing on record.
-On July 12th K2BWR worked G3GGN again.
-On July 13th K2BWR and W3RGQ worked G3GGN
-On Sunday morning W1BB worked G3JVI at 0325GMT/Heard G3GGN weakly.
-[illegible] W1BB's signals were heard regularly EVERY Sunday morning for EIGHT weeks at Strengths from 349 to 589
-G signals were heard and worked only on 3 mornings. Faint signals on two other occasiona.
-The high static level kept the weak 10 watt Gs from coming thru.
-Sunrise time - G land , as in the winter, was found the ideal time for signals to peak and make QSOs

1955 SEP-DEC -This "SEASON" instead of "Tests" every other week end or so, it was
1956 JAN-APR decided in cooperation with ShortWaveMagazine and W1BB to have the "Tests" EVERY sunday morning - then, if the in-between Sunday happened to be an excellent one for DX it wouldnt be missed. This through December to March 1956.
-During SEP-November, VS6CQ,VS6CW,VS6CZ ran special tests with W8GDQ and W1BB to try and see if QSO could be made. While these schedules were kept regularly every week, no contact or signals were heard. VS6CQ "sparking" these tests has now gone back to G.

1955-1956 SEASON - SUMMARY of this seasons OUTSTANDING events, follows, and is noteworthy for the large number of FIRSTS!!!

#1 -"FIRST!!" - DL/W QSO when "Shely", W3RGQ worked DL1FF, (Who secured special permission to operate on 160 from Dec. thru Feb- December 8th at 0510 GMT!! -- Followed by W3FBV, W8GDQ, W1BB same morning -- DL1FF was on whole season, with other DLs.

#2 -"FIRST!!" - YN/W QSO, December 15th when W2QHH worked YN1AA, (1oo Watts) for FIRST Phone/CW QSO, on 1805 Kcs. YN was #59.

#3 -ZL1AH worked the long distanc e to G6CJ three times, on Oct. 13,14, &15 - [redacted] S igs weak.

#4 -"FIRST!!" - HE/W Qso on February 12th when [redacted] HB1CM/HE, Liechtenstein, (HB9CM on DxPEDITION) worked W1BB at 0505 GMT, after a previous unsuccessful attempt. HB1CM/HE's signals came thru repeatedly and unexpectedly at 0505-0600, rst 229-449. This was an unusual performanc e on HB9CM/HE's part, and he is to be congratulated. (See separate story, which has been written on this for the magazines, with photos.)

#5 -"FIRST!!" - W/VP2 160 meter QSO ever with TORTOLA, British Virgin Islands, LEEWARD Group, when Howey, W2QHH worked FP/8AK/VP2 (W2BBK) on 1801 at 4.40 AM EST March 5th 1956 to receive 449 just as he was packing up to conclude DxPEDITION.

#6 -"FIRST!!" - Barbados to USA ever QSO, April 6th 1956, when VP6RG, "Ron", a high school Boy, snagged W1BB on April 6th for a report of 429x, BB being 569x. This represented a lot of hard work, perseverance and good planning on VP6RG's part, Of special interest is the fact that the pole supporting one end of the long wire antenna was our in the sea, cuite a fe w feet from shore so that quite a part of the antenna was OVER salt sea water -- NICE !!!

#7 -"FIRST!!" - Recorded USA/CUBA Qso on 160. W1BB and CO2BL worked it May 31, 1956 at 0700 GMT through heavy QRN. BB,559 -BL,449. "Doc" very cleverly and quickly "Converted" an 40 meter antenna for use on 160 - and it worked out -- well, well enough!!!

#8 -W1BB's signals were hea rd, for FIRST SWL report - by E rnest S prock, in SURINAM, PJ2CK, South America, on July 1 1956. This is interesting news from a Country rarely heard from.

#9 -An unusual "FIRST" was made, when after a series of schedules, the team of OM/XYL, ZS2GE/ZS2KZ got across to W1BB on July 11th at 0505 GMT, also S eptember 8th 0400-0440 GMT for the very "FIRST" ZS/W Qso on 160. (A special SUMMER BULLETIN to [redacted] maga zines, covered this event in detail)

#10 -Another important, and RARE, "FIRST" !! was made by CP5EQ, Bolivia, S.A. putting that country on the 160 Dx map, by working W1BB, July 22nd at 0800 GMT to receive RST 119-349 on QSB, with bb at 449

#11 -"FIRST!!" - British Guiana/W Qso by "DAD", VP3AD 119/339 -W1BB the same, on Sept. 12th. The previous week when contact was [redacted] about to be established "Dad"s xyl announced the expected Jr. Opr. would arrive, and the schedule was QRT just in time, but fast!! - A fine little fellow did arrive!! Congratulations!!!

- These summer FIRSTS by W1BB, are the result of testing out his theory that when it is Summer up North, it is WINTER down South, and conditions would be best for the Southern Countries & it worked out that way - but meant a lot of "EAR DRUM BUSTING" by static, for W1BB to keep these schedules.

- In general it was thought that the EAST/WEST conditions were not as good this year as last - believed due to sun-spot cycle - However, as will be noted - a lot of DX WAS worked, and some have the feeling that 160 meters is the least affected band of all.

- PARTICIPATION was good --Totaling 16/1s,24/2s,16/3s,17/4s,6/5s, 3/6s,2/7s,23/8s,15/9s,8/øs - one each VE1,2,3. --- Dx as follows: Gs-26/1s,3/2s,5/5s,4/6s,1/8 --- 3/KZ,4/KP,2/TI,2/GM,2/HB,3/ZL, 3/VS,1/EI,6/VP,1/FL,3/GI,1/OA,9/DL,3/XE,2/YN,5/OK,1/GD,2/KV,1/SP, 1/HR,2/GC,1/HE,1/KH,1/CO/2/ZS,1/CP and one VP/3.

W5CGT [redacted]

- WORKING Dx were W/VEs,,, W1BB,AHX,EPE,BMW-W2EQS,QHH,K2BWR/W3RGQ, EIS,FBV,TBG,DGM,MSK-W9PNE-W8IL---HEARD only by Dx - K2JIO. -----

- WORKING W/VEs were - G3PU,GGN,KKP,IGW,JEQ,ERN,JVI,FPQ-G5RI,JU,CV,NS G6GM-G8JR-KZ5PB,DE,FA-KP1SD,KP,CC ,DH-VP7NG,4LZ,9BM,6RG,3AD-DL1FF,DA-XE1A,OK-YN1AA-GD3UB-KV4AA-HR3HH-KH6IJ-ZS2GE,KZ-CP5EC-CO2BL --- HEARD only by W/VE were G3HWI,IL0-TI2BX-EI9J

PY1-15652/Rolf:

MAR 11 1962

Here is all the latest on 160

Thanks for al your fine assistance this year

If you can help me line up a PY HAM that will QSO me

Or make private schedules - would much appreciate

Happy summer and 73 de Stew/W1BB

ROLF - I am enclosing some PUMP literature that may be of interest - apparently you are in same line of business that I am - eh??? CLENZ s/a BOMBA!!!

Also am sending you our 1962 calendar via slow mail

Stew

W1BB
STEWART S. PERRY
36 Pleasant Street
Winthrop, Mass.

Please be sure not to forget to line up some HAM who will make schedules with me on 160 - I might even send small 160 Tx if necessary if can find reliable fellow - most so far havent been reliable - did not keep schedules etc and only wan wanted me to send them somethings.

FEBRUARY 28 1962 W1BB -- 160 METER DX -- BULLETIN 1961/1962 - No. 4

de W1BB/General Remarks --- Well!!! The last regular TRANSATLANTIC DX TEST has co and gone, and the CQ-CW-WW Contest is over - so no doubt DX activity on 160 will begin to slack off - as spring approaches!! --- 1962/1961 Season hasnt been as go as some expected, or as bad as it could have been. This comparison, for example, total number of different DX stations worked by W1BB each year, might be of inter however. 54/55-19, 55/56-26, 56/57-20, 58/59-4, 59/60-5, 60/61-11, 61/62-27!!! I indicates a definite cycle, corresponding with Sun-Spot activity, and improving!! --- While 160 DX isnt completely done for this season - more DX will be worked - apparently we have had about the best of it for this season - or have we???????? --- Dont forget the new "Format" wherein underlined calls indicate QSOs, without underline, indicates "Heard" only. ---- The reports from participating stations, SWLers are just grand!! Keep 'em up and TKS!!! ----- Reference IQSY - it has not yet been determined if 160 meters data which might be collected, would add enough knowledge to Ionospheric conditions to make the venture worth-while, as the 50Mc Data was for IGY -- More on this later. --- There will probably be one more Bulle this season. ----- Occasionally an error, sometimes due to W1BB fault, and at oth times not - creeps into this bulletin --- Please excuse the "Human-Element" O.M.s However on the whole, information herein is believed to be 99 and 99/100 % Pure!!

W1BB/Summary LOG of SUNDAY MORNING 160 METER DX Activities:::

1. February 4th 1962, FIFTH Scheduled TRANSATLANTIC Dx TEST --GENER

At 0000z DHJ54/2 1800k was coming thru 549, a good omen for Dx, At 0500 449. Quit a lot of Snow Static - Wx Cloudy - Band opened up to East, Fair to good - South a W[illegible], Poor. ------- EAST -Fairly good Dx - Gs Coming thru at 0530 peaking 0730, t loudest signals lingered until 0820. Gs,5PU,6BQ,3ERN,CHN,5JU got across. Exciteme of the evening was new Country heard from when SVØWZ reported thru G3PU he was Hr W1BB 0505-0570 - unfortunately no QSO resulted. Reports as follows --G/SWL/Allen: VP8GQ,1BB,PPN,TX,2FYT,GGL,KHT,8EVL,K8EVG-VE1ZZ-ZC4BP-8HBR/8 - Thanks Geo!!! Then G/SWL/Smith: :VP5FH,1BB/1,HR3HH,2FYT,8NSF,1PPN,2EQS,VE2AYY-UB5WF--VFB Norm & Tnx!!

G3PU/ERN:: VE3BQL/SU/1803, SVØWZ, 1BB/1, PPN-2FYT, IU, KHT-K2PNF-K8EVG, HBR/8-VE1ZZ, VE2AYY, -HR3HH ----- G2PL/Peter:: 1BB/1-HR3HH ----- "DUD"/G8CJ, of "Shadow-Edge" 160 Dx fame, is back on 160:: VE1ZZ-1BB/1, PPN-2FYT, KHT. Says friend ZL1AH getting "Hot" on 160 for QSO w/HC1AGI & Others ----- G/SWL/IBALL:: 1BB/1 ----- G/SWL/Davids BB/1-VE1ZZ ----- VE2AYY/MIKE:: G5JU-VP5BP. Could have wkd VP5BP except he was to close to 1800k & BC Harmonic - suggests ALL Dxpeditions and DX stay away from ev multipliers of 10kc, 1800, 1810, 1820 etc - in future!! ------- WEST:: Conditions po A few 6s hrd weakly - no real DX. K7HDB/Dick active, but no Dx. ------- SOUTH:: HR3H was star of the evening QSOing quite a few Ws. Real excitement, when HR3HH report thru W1BB he was hearing G3PU - BB Notified PU - tried get them together - but n QSO resulted. This is first G ever hrd by Hal/HR3HH --- HC1AGI came on at 0445 CQi many times - many Ws called him-few QSOs-apparently skip conditions poor and sta bad --- A W was hrd clg VP2VL --- XE2OK also on --- BRAZILIAN/SWL/ROLF/ISWL/PY92 :: WØGBV, 559/239-K8BBI, 349 - VFB & Tnx!!! New Country hrd from on 160!!! VP8GC:: W6KIP, ØCDP, 5SOT all Hrd VP2ZA ------- NOW HEAR THIS de ZL1AH/John!!!-"Stew, I am gradually readjusting my ideas again as the sun-spot cycle takes a dive. I was t this week that you had a test this weekend so took a couple of looks at the band. At 0809 hrd W1BB/1 on 1807kc calling CQ-DX until 0810 RST 339 - would have been better except for summer static" VFB&Tnx John-sure glad you are back on 160!! We all be looking for you!! ------- HC1AGI/"Don"/W3EIS :: W8NSF, GIY, FW, WI-K8BBI -- seems like the 8's had it!!! ------- OTHER PARTICIPANTS - W6KIP, 8GDQ, HUT-9JDK-ØC VX0/9-VE1DR DX--G2IM-G3CHN, PEK, ORD, LI, OQT, NXJ, MZW, NXV, GGS, BJD, GXP, NMF, NEE-5JU- ZC4BP-OK1GT-GW3CBY, PZU-GW8PG-GI3NZZ. --- HC1AGI says - "Conditions Lousy" !!!

2. February 11th 1962, Non/Scheduled SUNDAY MORNING 160 Meter DX GENERAL --- Snow Static QRN/S7 at 0430 as snow storm passed thru Eastern USA, cu down Dx on what otherwise might have been a real good AM - all in all this was a rather poor Dx morning. ------- EAST - No Dx hrd at W1BB - Marker DHJ54/2 not hr G/SWL/Allen:: 1BB/349-2FYT/239-CN8PZ-ZC4BP-SVØWZ-EL4A-UB5WF -Tnx Geo, --- GSWL/Sm 1BB/1-2FYT-CN8PZ-Tnx Norm!! ---GSWL/Davison:: 1BB/1-CN8PZ QSOing G6BQ @ 0516/1805 TH --- HR3HH QSO'd W1BB/1 altho going was ruff w/S7 QRN!! --- HC1AGI/Don, rep "QRN so heacy all reception hopeless"!!! --- VP8G, apparently QSO'd W3FBV & W2FY No confirmation, --- W9PNE/Brice:: VP8G/1801/339-HR3HH ---HC1AGI did hear HR3HH VP3AD/"DAD" Daly was on and QSO'd VP2VL589 Tortolla Island, (Box #45 P.O.)-"Dad" and VL were both real excited for their FIRST EVER VP3-VP2 160 meter QSO!!! Con to both - "Dad" thinks 160 good for another month at least. He also hrd, W1BB/1- 2FYT-K8HBR all 569 --- PY/SWL/Rolf-NITEROI BRAZIL!!! Hrd VP2VL/449 & VP3AD/239 sure nice to be getting SWL reports from Brazil - not to get a 160 meter Tx goin there!!! Tnx Rolf!! ---WEST - W6KIP/349 QSO'd ZL3RB/139 @ 0800 - CERY FB!!!!!

3. FEBRUARY 18th 1962. - SIXTH ANNUAL TRANSATLANTIC DX TEST GENERAL - Quite disappointingly poor - of all Mornings - on the last, 6th Annual Transatlantic DX TEST - and also the ARRL Dx Contest to have the QRN "Gremlins" work spoiling things!!!!! SKIP, all directions, pretty good - however, S6 to 8 QRN covered most DX stations with a heavy blanket - nevertheless, a few lucky St did succeed in breaking thru for QSOs. ------- NORTH - VE1RAN/FP8BX DxPediti expected on - did not show. ------- EAST - The European Dx conducted a battle ro with QRN, pushing thru occasionally to QSO on "Peak" conditions. G3PU:VE1ZZ -- G6BQ:VE1ZZ - G3FPQ:1PPN -G6GM:1BB/1 - [redacted] - W3EIV:G6BQ - W1BB/1:G6GM, BQ, G3FPQ - VE1ZZ:G3OIT and others --- GSWL/Allen: W1BB/1, PPN-K1DSV-2 2EQS, KHT-3EIV-K8OR-VE1ZZ-VO1FB-HC1AGI-VE3BQL/SU-SVØWZ - Fine list Geo!!!/Tnx!! - GSWL/Smith; 1BB/1-2KHT-VP5BP & other unidentifiable signals Tnx/Norm!! --- K4KSY/Drayton hrd only DHJ54/2 weakly at 0330 - QRN S7/8 --- GM3IAA/Jim MacIntos Hrd W1BB/1 - says "It is strange why our Northern stations do not get across so on 160!!" Their strictly low QRP has something to do with it no doubt!!! Jim use to operate a TELEFUNKEN 5 KW Spark Tx at Assiut Egypt - some difference!!! GSWL ISON:1BB/1/569-1PPN/459-2FYT/449-VE1ZZ/559 - Tnx Henry!! ------- SOUTH - VE3C Noel Eaton, Vacationing Canadian Div. ARRL. Director broke thru as VP5BP, Grand Cayman Island Working W1BB/1, 3RFA, 4YHD, 8HGW, and probably others for NEW COUN QSOs. (On Saturday Night 17th QSO'd W1PPN). --- DON/HC1AGI/W3EIS, apparently was having QRN problems too. Many W/Ve frantically and frustratingly calling, 0730-0 w/no Ans. However he did QSO W1BB/1, 2IU, MSK-4KFC, YHD-6KIP-K6ZH-9PNE-KH6IJ - VFB WEST - Practically NO Western Dx heard at all except W6KIP and only for brief t

ADDITIONAL PARTICIPANTS - W/VE: 2BWR,SSC,GGL,QHH-3GQF,CJ-4OMW-K4KSY-8GDQ,TSD-K8[illegible]/8,IQC,RRH ---Dx: VP8GQ-GI6TK-GM3NHO-GM3IAA.

FEBRUARY 25th 1962 - CQ-160-WW-CW-CONTEST --- GENERAL - Hoped for A[illegible] Contest/Dx conditions did not materialize - principally due to high static levels. Mu[illegible] Dx was in there but could not be copied or worked thru QRN - although some, w[illegible] All in All, a good TEST was had however - everyone seemed to have a grand time, an[illegible] a lot of new stations were heard from!! CQ will carry the results. W/VE - W1TX/Roy FP8BX,WØVEH/VP9-HC1AGI-EI9J-HR3HH & others for score of 10,430 in 11 hrs -Not bad- had bad 60 cycle line noise!! ---W1BB/Stew: VO1FB-WØVEH/VP9-FP8BX-HR3HH-HC1AGI-G3[illegible] GD3UB.--- W2UKS/Bill - KH6IJ-EI9J-G6BQ-WØVEH/VP9-HR3HH-GD3UB-XE2OK-FP8BX - Tnx Bi[illegible] (Not QSO'd) ---VE2AYY/Mike:-"Had a ball"!! 152 QSOs including FP8BX-EI9J-HR3HH-WØVEH/VP9- now has 10 Countries on 160 - says Cdx poor to West. Hrd XE2OK,GD3UB,G3[illegible] Score 11,760 ----- Willie/W8GDQ - had some fun, but snow static terrific - missed few good ones. ------- EAST - A LOT of Europeans were on for the test, but between QRMing each other, and covering up DX, they had a lot of trouble with the QRN also Gswl/Allen:W1BB,FRR,MR,PPN,TX-2EQS,FYT,IU-K2BWR,DGT-3ECR,EIV-4KFC,YHD-8GDQ,HGW-VE[illegible] VE2AYY,3KEE-VO1FB-WØVEH/VP9-HR3HH-EL4A QSOing Gs & OKs but no Ws - Nice going Geor G3PU/ERN,says "Without doubt, best opening of season BUT ruined over here by lot Gs[illegible] in Contest calling "CQ-DX" on top of W/VEs - they were even on W1BB's Fy ev when you yourself were calling CQ or working another station - I cant understand some of these chaps!! On top of that, W stns did not seem to be tuning our Fys - I packed up after 50 Mins fruitless calling!!! Ern heard: W1BB-2FYT,HUG,IU-K2DGT-4KF 4YHD-8GDQ,HGW,NSF-K8HBR/8-WØVEH/VP9-VE1ZZ,2AYY-VO1FB-HR3HH-FP8BX!!! --- G3CHN/Roge W1BB,ME,PPN,FRR-2FYT,3GQF-4KFC,YHD-EL4A-EP2BK-OK1ZL,NR-UB5WF-VE1JX-HB9T-ZC4PB-VO1[illegible] WØVEH/VP9 589 at times!!! and many others hrd - however "T"was a bit of a carve-up at times" he says but very enjoyable, bringing in no less than 4 new ones!!! K8HBR HC1AGI-HR3HH-WØVEH/VP9-KH6IJ-XE2OK. Score 12,792 for 152 QSOs - 3,Dx and 39 Mult. Hrd no Eu in FEB - Al leaves for his job - sailing on the Great Lakes, shortly - T loaded ship antenna that he appropriates for 160 DX goes with him - best luck Al!! Gsw[illegible]MITH;W1BB/1,AWP-2FYT,UWD,IU,GGL-K2DEFT-4UK-8HGW-K8HBR/8-VE1ZZ,2AYY-9PNE-WØVEH Nor[illegible] also sent in FB list for 24th also - tnx & Congrats!! ---- CYPRUS HEARD FROM AGAIN ZC4PB/GERRY BATEMAN - With low power and poor antenna - single wire 15-20' H made 8 Eu. QSOs Hrd Between 0222-0430z the following: WØVEH/VP9,W1BB,PPN,ME-4YHD-3 3GQF-8HGW-K8HBR-No VEs hrd. Stan called Ws using 1827Kc but ND - Unfortunately - T a million Stan!!! ----W1PPN/Cliffe: G3CHN-GD3UB-G3IGW-EI9J. ------ WEST - W6YC/Gen using 235 ft LW on small 30x120 lot QSO'd 25W States, VE,KH6 & XE1 - heard W1 & VE and also HC1AGI/579 but could not QSO him!! ---- NORTH - Nil!!! ----- SOUTH - HC1A[illegible] :1BB,TX-K2PNF,DGT-3GQF,EIV-4AM-8GIY,GDQ-K8HBR/[illegible]-9YT-ØGBV-VE3KE-XE1OK-HR3HH-YN1AA. "Don" says his problem hearing DX - his European listeners mail is terrific - but he doesnt hear 'em - "Two whole nights without sleep - wonderful conditions - but only 39 QSOs!! Don corrects BULL No.3 - HC1AGI is 11,700 Ft high - not 1170!! --- OOOOoooooPs!!! Sorry!! (Don also had nice string of 23 stations QSO'd on 24th) --- NEW COUNTRY HEARD FROM - PJ2AE/Stan - Who says he cant operate on 160 - regs - but thought we would be interested in signals hrd, as follows:: W1BB/1,PPN,ME,MMM-2FYT NVD,IU-K2DGT-3WV,GQF-4YHD-8HGW-K8RRH,MBR/8-9YT-WØVEH/VP9-ØBDS. We certainly are interested and Tnx millions Stan - and get after those PJ Authorities to allow 160 ADDITIONAL PARTICIPANTS None not already mentioned!!!

MISCELLANEOUS TOP-BAND ITEMS ------- W6KIP/Alex QSO'd HR3HH & HC1AGI - Is disappoi[illegible] there is so little DX to West. Has "Rectification" Problems with his antenna tower causing QRN - says "Noise here remains tremendous problem and I dont know WHAT MMM I'd do without noise balancing effects of receiving antennas - although it cuts do on signals - it still gives much better signal to noise ratio, and I seem able to [illegible] all stuff fellows in QUIET spots hear"!! -------G8NF/FLASHBACK!! Jan 7th Harold Hr VE2AYY/57-VE3KE/57-VE1ZZ/57-W1BB/57-HC1AGE/33-W2UWD/45-K3KMO/45-W8I[illegible]/33-3NTX/33-1C[illegible] - Tnx Hal!! -------MM GM3IAA/JIM, A real O.T. -1914-1915 R.E. Wireless Ass' [illegible] O.T. - Heard Jan 7th 1BB/1-569,1PPN/349,1TX/349. He is Ex VS1AA-2AA,2AF. In 1934 improved WINDON antenna which is written up in RSGB handbook -- He is looking for Contacts!! Nice to have you with us Jim - Good Luck!!! (Has just wkd 1BB & 1PPN!!! ------- BRAZILIAN/SWL-ISWL/PY9247-Rolf Rasp --Checks in with Vy nice ltr & fm now [illegible] log 160 stations and QSL - has been HB9SWL 5 yrs - also DL,PK1,PK3 - has conv [illegible] to 10 tube BC/Rx Home Made. Sure nice to hear fm PY - Tks!!! ------W-VE3DU/Dave [illegible] on 1898 Kc every Sunday 0600 - some quite [illegible]

VØAIH/VE3 - Loran does not bother there (Lucky boys!) However TV puts birdies 1825 1827 messing up Fm Dx-Thinks Cdx not nearly as good this year as last (Dont unders that dave!!!) Jan22:G3CHN,HR3HH,HC1AGI - Tnx Dave!! ------- W6KIP/Alex sez Europea who want to work W5,6,7, MUST use 1800-1805Kc - hopes for 160 in KL7 someday - Try fo[illegible]LAD & VOØLL on 160? ------- K7HDB,JAN27:2FYT,UWD rapid QSB -Jan 28 1BB/1 --- Does anyone have a "Yen" for WHALE MEAT??? write Dick!! ------- W2TR/Geo -"We dont Hv much time for HAM wk, but when we do, it is the OLD TOP BAND, although things a rough at times, we enjoy 160 meters best of all!!! -------K4KSY/Drayton - "Notice that West Coast stations come in hr quite early of late - sigs break thru literall everything is quiet on top band - then suddenly they appear-early as 2030est-sever nights exceptionally good 05 low as 1980kc - Alas!! cant seem to wk 'em!! Does thi fit with study E/W Cdx?? ----- WE ARE SORRY that one of our most reliable & well K 160 DXers BRICE/W9PNE is plagued with ulcer trouble requiring strict diet - but wo of all enforced sleep schedules that cut out much 160 Dx activities - better obey "DR's orders" tho - get it over with - and then BACK TO 160!! Best luck Brice Fm a of us. ------- WØCDP/BING says skeds w/ G3PU,6HB-EL4A-DL1FF on 1999 did not wk out Hrs VP8GQ well on 1827 - HC1AGI best on 1995 ------- OH3NY/Matti - Continues effor to get FINLAND Govt allow 160 agn - Keep up the good wk Matti!!! ------- HC1AGI/Do rpts FB string QSOs Jan 14 & 28th --14th:1BB,ME,2KQT,FYT,IU,WA2KVH,8GDQ,JIN,K8IJQ, HBR/8,K6ZH,KH6IJ,ØNWX,CDP,IFH --28th:1BB/1,2IU,FYT,UWD,GGL,K2LWR,3RFA,GOF,6KIP,8FG FW,IQSmLOF,HRV,NSF,HGW,K8HBR/8,ØEUQ,CDP,IFH,VP8GQ,VE1ZZ,ZL3RB!!! VØAIH/VE3 ------- EL4A/Ken - reports via Ham Radio "Been hvg trble w/Kw rig but expect hv fixed soon -------K7ICW/A1 - has new Hy-Tower coil loaded Vert. 160 ant!!! Says DX poor this but QSO'd KH6IJ-line noise teriffic - going to hit CQ-WW-160-CT hard this yr agn!! -------W2UKS/QSO'd VP5BH/1804Kc Feb 19 - Hrd FP8BX/1827k,20th-sez QRN HIGH!! ----- SORRY TO HEAR?W8ANO/Waldo - has sold 160 meter Tx&Rx-gone SSB!!! He'll be back - s day!!! -------Willie/W8GDQ Moans that cdx havent been good as expected this yr & w power limit/line noise etc has had rough time - sorry!!!!

BOY oh BOY/W2UKS/Bill!!!!!!! Feb 21st truned in an incredible list Dx Hrd - UB5WF, [illegible]VP5AD-HC1AGI,HR3HH,G3CHN & Others ---W1BB sed to himself - there's something here worth knowing - SOOOOooooo wrote Bill, asking if by chance he had real QUIET location - and here's the answer of 160's UTOPIA!!! "I am Located on an island 2½ miles out in the Atlantic Ocean Fm mainland -plenty water all sides-plenty deep - noise level S2-S3 normal - in Loran S5 - we have NO industry in or nearby/not allo I'm using Heathkit "Q" Multiplier HD-11 w/SX28 & L.W. 200' long over large parking lot & 40' high. QTH center of city w/4&5 story hotels arnd - resort city/fairly qu in winter all directions except N - Ant runs E/W-end into shack - ant coupler-Viki II Tx/200w--sure can hear 'em/darned if can wk 'em!! Wl keep trying (And he DID!!) "Ed-Note" -W1BB also had superb reception when band was open one Sunday AM, and he was at Summer QTH 300 miles into Maine away from noises - hearing and working UB5, ZC, many others - Dx sounded like 20 meters - all of which goes to prove that the most important ingredient for 160 Dx is a quiet QTH in the clear.

PROPAGATION - We 160 Meter fellers are especially interested in this - and so for everyone's benefit we copy herewith all the latest dope - Courtesy "FLORIDA-SKIP"- "PROPAGATION CLEARED UP" -!!??- Propagation charts show an easing up of the rate a which Dx is easing off, This is proof of the slowing up of the slowdown. ---- It should be noted that the slowing up of the slowdown is not as good as an upturn of the downturn which is a good deal better than a speed-up of the slow-down. It does suggest that the propagation is about right for an adjustment of the readjustment. ---Turning to long skip, we find a definite decrease on the rate of the increase, which clearly shows a letting up of the letdown -----Of course, if the slowdown sh should speedup, the decrease in the rate of the increase of short-skip would turn into an increase in the rate of decrease of long skip. In other words, the decel/s would be accel/erated. ---The charts suggest a leveling off, followed by a gentle pi[illegible]up, rather than a faster pick-up, a slow down of the pickup, and finally a leveling off again of the pickup. --- At any rate THE PROPAGATION IS RIGHT FOR A PICK-UP this winter - that is, if you have a 10 Kw rig and a 25 Element beam!!!" forecaster Johnny Short Skip/W4SKIP. --PS - How's DX fellows??????? There you have it fellows - and this information should be VERY helpful in working 160 meter DX!!

73 de Stew/W1BB - and HAPPY HAMMING to you - Especially on 160!!!

APR 20 1962

PYswl/Rolf

Here is all the latest on 160 for you with compliments W1BB

Thanks for yourletters March 8 and 19th

The reports were very interesteing and studied carefully and entered in SUMMARY log here.

Do not know VP2VL QTH - suggest you write VP2VI W2YH T.W. WINTERNITZ, Yardley Road, Menham N.J. - He will know or DX Magazine, Don Chesser, Burlington Ky - they will surely know. Sorry I cant help more.

W1BB
STEWART S. PERRY
36 Pleasant Street
Winthrop, Mass.

I sincerely hope that you can get someone on 160 for schedules with me next fall - or send me likely suggestions to write to.

That SX71 is a real fine receiver - VFB - Know you will like it,

VERY VERY VERY FB on you and PY1NFC -- and with 40 watts if you can build it to cover 160 we can QSO - and make HISTORY - since there has never been a PY/W QSO on 160 meters - and if and when you are ready if you will make private schedules with me , you can win the W1BB CERTIFICATE for FIRST QSOs - I will give you all the needed information on Antennas - and help you in any way possible - just let me know what you need - as long as you will make the skeds with W1BB only until a FIRST QSO. And I assure you that I have a sustem that will work if you are faithful and persistent!!!

HAPPY HAMMING and 73 de Stew/W1BB

APRIL 16 1962 W1BB -- 160 METER DX -- BULLETIN 1961/1962 - No. 5

de W1BB/General Remarks --- This last 1961/1962 season BULLETIN has been delayed a bit for unavoidable reasons, but the enclosed FLASH supplement was sent to all the Ham Magazines on time. This will be the last Bulletin until the 1962/1963 season opens this fall - unless by chance there may be enugh activity to make a summer Bulletin necessary, ------- Yes - it HAS been a GOOD season, a VERY good season on 160!! with indications of possibly better next!! We say "Season" because during the winter months there is more activity and Dx worked in this latitude - the band is quieter, nights are longer, and weather males it more conducive to staying indoors at the rig. HOWEVER, we must not forget that while it is summer here, it is "Winter" elsewhere and there could be Dx minded Hams willing to work 160. In fact, W1BB wkd ZS in August, and CP in July. Most of us Northern Hemisphere people are so happy to enjoy our outdoor activities in the warm summer months, that our 160 Dx, generally speaking will be QRX until about Oct or Nov. Another advantage however, these periods of activity and rest give more ZEST to the Test when it resumes!!! SOOOOO generally speaking QSK till fall and b.c.n.u. then. ------- We cannot end this season, (or any other), without saying many many thanks and appreciation to the SWLs whose prompt and complete reports have added so much to our 160 meter tests. Outstanding, as usual, were reports from G/SWLs, Robert Iball-Norman Smith-George Allen,--Excellent reports from SWLs - Gs/John Hall, Henry Davison,Jim Sleight -VK Charles Thorpe- PY/Rolf Rasp -- not to mention a number of others who sent in spot individual reports. Next year, it is hoped to make available to any W/VE Hams who are interested and write for them, copies of these reports. "Cheers" to these bo - and fellows,keep up the good work!!! -------W1BB wishes to send thanks and 73 to all the 160 meter gang with appreciation for their QSOs, reports, cooperati and all matters which have led to greater enjoyment of DX hunting and testing on

W1BB/Summary LOG of SUNDAY MORNING 160 Meter DX activities!!!!

MARCH 4th 1962, Non/Scheduled SUNDAY MORNING 160 Meter Dx --- Genera speaking this was a dandy morning for Dx to Europe - Stations "in" for 2½ hours!!

W/Ve-REPORTS:-W1BB QSO'd 11 stations!! GM3IAA-G6GM,BQ-G3OIT,IGW,PJA,GWX,PU,FPQ,E
OQT - some up to rst589!! ---W1PPN Topped W1BB with 13!!! GM3IAA,WØVEH/VP9,A3-G3R
OQT,KOX,PU,ERN,FPQ,IGW,OIT,MBN-G6BQ- also AM/SSB QSO with G3FPQ - Cliffe on AM
Congrats Cliffe!!! --- VE1JX:: -G5JU,3CHN,IGW,FPQ --- W8GDQ/Willie::G5PU,W6KIP,"H
Eastern guys working stuff but couldnt pull 'em through"!!! ---W2UKS/Bill- Hrd On
W6KIP,WØVEH/VP9,G3PU,G6BQ ---WØAIH/VE3 QSO'd G6BQ -------(EAST) - VE3DU/Dave - G3CH
PU,FPQ,6BQ says "Good opening but plagued with S6 intermittant line noise and coul
NOT make QSO through it"-Sorry Dave!! -----G/swl/Smith::1BB/1,PPN-2GGL,FYT,IU,UWD,
KHT-8NSF-K8HBR/8OVE2AYY-VE1JX-"No exciting stuff,a W did call a VP" -Tnx Norm!! -
Gswl/Allen::says "A very good time was had by all-best condx since I came back on
trail!! Spent much time on 1.9-2.0Mc hrd nothing-though pptomistic-Willie/8GDQ do
have the punch of old-wonder if his VERTICAL deteiorated?? -(NO!! Power limit now
50w instead of 200 his area) - anyway real nice to hear him around again" -Hrd, 1
1/589,AW/349,TX/449,PPN/559-2IU/339,FYT/559,UKS/449,GGL/349,TR/459,KHT/349,UWD/45
K2BWR/229,K2HLI/239-3NOQ/239,GQF/239-K5TMF/559-8GDQ/469,NSF/339,FW/358-K8HBR/8/33
VE1JX/559,2AYY/449-WØVEH/VP9/239 - Tnx Geo!!! -----G3OJI/Jim had an exceptionally
list 1BB/1,PPN-VE1JX-K8HBR/8,8NSF-VE2AYY-W2IU,FYT,8FW,K2HLI,2UWD TNX!! ----Gswl/I
1BB/1,PPN,2FYT,VE1JX -----G5PU/Ern - "Without doubt best morning of season"!! 1BB/
PPN,2EQS,FYT,GGL,IU,UWD-8GIY-K8HBR/8,8NSF,WQV,VE1JX,VE2AYY,VE3ABG -----Gswl/David
W1BB/1/569-VE1JX/569-W1TX/449-2FYT/239-VE2AYY/469-1PPN/459-8NSF/339 - and proud o
his FIRST W8!!! -----G3OQT/Dick - Hrd 1PPN QSOing/569-WØVEH/VP9/439 CW then change
to phone hrd Cliffe 3/5 and VP9 2/3 - also 1BB,AW,2FYT,IU,8GDQ,1PPN, VE1JX,VE2AYY
WØVEH/VP9. ------(WEST):Nil ------(NORTH):Nil ------(SOUTH) - ZL3RB reports,"condx
unusual, extreme slow QSB,sigs rising to medium peak slowly, dying right down thru
period of 5 minutes. ZL3QX hrd W6HRG/459 -We logged W6KIP,5SOT,6HRG,ØCDP. It gave
ZL gang real thrill hearing Stateside clg "CQ-ZL" and we really appreciate the Ws
looking for us. Participating were ZL3QX,ZL3BC,ZL3RE,ZL3CO and ZL3RB"- Tnx ZLs!!!
Another participant this morning was XE2QK.

MARCH 11th 1962 - Non/Scheduled SUNDAY MORNING 160 meter Dx --- Early
skip condx seemed fairly good and the band did open up fairly well. -----(EAST)-1
hrd EP2BK briefly 119-239 0310z,later0335z G6BQ came thru 579 QSOing W1BB. At 0500
few other Gs hrd,G3PU coming thru real well. WØ/AIH/VP9,QSO'd a number of Ws ---
G3OJI/Jim-"Cdx rather poor-Pkease ask W2FYT NOT to send QSX 1801kc as many Gs sta
calling there and make mess of Dx band Hr - HC1AGI/VFB!! Hrd1BB/569,HC1AGI/559.3G
439,2FYT/569,1PPN/339 --- G3FPQ/Hrd HR3HH ---Gswl/Smith:1BB/1,2FYT,IU,3GQF --- de
G3PU/Ern:1BB,PPN,2EQS,FYT,IU,3GQF --- G/swl/Anthony Quest;1BB/449-K2DGT/339-3GQF/3
W9PNE/Brice:: says his son had emergency appendectomy this night so vy QRL,however
did hr G5PU/229. Brice feeling better himself - using NAVIGATOR/50w. CQ/CW Contest
105x38 ---Gswl/Allen:1BB/1/349-2UKS/229-4KFC/229-K2DGT/349-3GQF/459-2EQS/239 ----
GM3IAA/Jim:1BB/1,PPN,3GQF all 349 - Jim is planning new antenna for next year !!!
WEST;;;; No Western Dx hrd hr or being wkd by others -------(SOUTH:);::: Real excit
ment when W1BB alerted Gs to fact that HC1AGI was on 1826 - thereupon G6BQ QSO'd
for FIRST!!!. G/HC 160 Meter QSO ever. HC1AGI gave G6BQ 449. Congratulations!!!!!!
In addition HC1AGI QSO'd a vfb string K8HBR/8,3GQF,2EQS,This one the reward of ma
hours of trying and a new antenna by Chas -2GGL,K8HUT,2UWD,1PPN. Don hrd & Cld G3
several times but he did not answer. Interesting that G6BQ who was alerted could
Hc and G3PU also alerted at same time could not - proves that 160 meter skip can
be either "in" or "Out" because of short distances separation between receiving Q
HC1AGI hrd HR3HH/579 and W1BB/579. ----- HR3HH had Fb sig QSOing W1PPN,2GGL and m
other Ws -----W9PNE;HC1AGI/449,HR3HH/569 -----(Other PARTICIPANTS) - Ws,1TVN,2DGT
TR,UKS,4KFC,MUTk/1,8HUTk,GDQ,HBR/8,TYSk,FW,GDQ,ATPwa,VLC,ØTUT --G3ERN,LIQ,PGN,GM3
OKINR, SWL/Allen,Smith.

MARCH 18th 1962 - Non/Scheduled SUNDAY MORNING 160 Meter Dx ---(GENER
W1BB had to go to Maine on a business trip/spent one night at old Homestead 300M
N. Boston in quiet QTH - took SX100 and Ranger/50w along. Too busy to spend muc
time on air but did QSO EI9J and Hrd G3PU,ERN-6BQ-OK1ZL - what impressed most, wa
the quietness of reception. At W1BB's home QTH Boston, the noise level reads arou
250 to 300 usually, while at this quiet place in Maine it was only 140 to 160 - o
about HALF and DX signals sounded just GRAND!!! Imagine what a difference this
makes in weak signal reception - to get away from the Electronic Clouds that
surround big cities and communities, just like the "Smog" in California - or Smok
and haze -same idea. Ph for these Island and remoht QTHs for 160 meter Dx!!!!

(EAST) - EI9J:W1BB/1ØMaine -G3ERN QSO'd UO5AA - Other QSOs were made. ---G3PU/Ern "Heard only WØ I have hrd this season WØNWX,1802kc 0610z rst229/459. Also hrd weak WØon 1999 believe WØCDP. Records!!";;;1PPN,TX,2EQS,2IU,UKS,UWD-3ECR,GQF,ØNWX - VF W2UKS/Bill::G3PU,XE2RGC,HC1AGI & QSO'd W5SOT!! ----- Gswl/Davison::1PPM/459, thru se[illegible]e QRN ----- Gswl/Allen::K2DGT/339-ØVEH/VP9/349-K2BDM/339-1PPN/449-3CQF/449-2IU/329-2UKS/339 -----Gswl/Smith::2DGTk,1PPN,TX-3MSK,ECR,GQF-2GGL-IU -----(SOUTH)- VP3AD/Dad hrd 1PPN/449 0432 and W3GQF/449 0450 other signals in there but QSB and static made copy difficult so closed up early "Guess 160 has had it for this seas Tnx Dad and b.c.n.u. Next for sure!!! -----(WEST)-----(NORTH)-----(Other PARTICIPA Gs3OIT-GI6TK,VO1FB -W/VEs: K1TKK,8HRG,8UDR,BHUk,ØNWX,TUT,LQTk ---SWL/Davison,Aller Smith. Activity and Participation dropping off as Spring/Summer approaches!!!

MARCH 24/25th 1962 Non/Scheduled SUNDAY MORNING 160 Meter DX Activitie (GENERAL) - Dx participants dropping off - QRN very high due to snow/rain static - skip generally quite good -W1BB QRT at 0555 -----(EAST)- W1BB got on early again schedule with EP2BK - hrd each other but no QSO. Also made contact March 22 and 2 0220z approx. Static very heavy. W3GQF and W2EQS tried for Bob/EP2BK also. W1BB h him 119/229 to 000. W3GQF hrd him 349 W2EQS could not copy. Skip not as good as n before when W1BB not solid QSO resulted. QRN Vy high also - at 0320 G6BQ came thru with CQ QSOing W1BB,his 449/BB559-his signals at 0400 still 449 QSBing in and out-QRN high. W2KUT also took part DXing QSO'd G6BQ/Jack getting 349/449 and giving 4 W2IU QSO'd G6BQ getting 33/49 Giving 469 -----Gswl/Norm Smith:: 1BB/1,AW-2KUT,IU, K2PNF-3FBV,GQF says "One or two weak unreadable Ws including a W5 at 0517"!! Norm going to slack off for the season - thanks for some wonderful reporting and assist Norm OM. -----W2UKS rpts no DX hrd "Appears that springtime is here" -----(SOUTH)- (WEST) -(NORTH) -Other PARTICIPANTS) -2PSQ,4WWU,8CJN,9ZQOk,ØCDP-G3ERN-SWL/Allen & Smit

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COASTAL COMPARISON NET -(West) ----- K7HDB/Dick writes "There is a tremendous diff rence between signal levels at the Northern Vs Southern stations and I would like get some sort of NET of Coastal stations going to compare reports of signal level during these DX tests in an effort to determine the Southern ends of this zone of absorption or whatever it is that is creating the relatively poor Cdx at the more Northerly stations. Mainly interested in stations close to shore where Loran is li limiting factor in reception rather than power line noise" Those interested, write Dick - sounds like VFB idea!!!

INTERESTING !!! W1EAE/Bigelow - of EMN CW traffic NET says that on 80CW with lower Sunspot Numbers, LONG skip, and short Ground wave are setting in to such an exten on 80 that it is virtually impossible to conduct their normal NET. Will have 8 to stations on and NET Control can only hear 2 or 3 - they are eyeing 160 for possibl solution to their traffic NET Meetings 6.30-7.30 PM EST. Best of Luck Big and boys ZL STATIONS -W6KIP/Alex de ZL3RB/Mick, says more ZL activity. Look for ZLs 0720/0 z around 1975-2000Kc. ZL3BC/Joe,ZL3CO/Garth,ZL3QK/Jack & ZL3RE/Art - all active n Also ZL1AH. This one FOR THE RECORD!!!! "ZL" --FIRST !!! ZL/W QSO ever was betwee ZL1AH/John Eightman Dec 31st 1961 and W9CVQ - the Second between W1BB and ZL1AH A 16th 1952. ZL1AG has two other FIRSTS!! ZL/G and ZL/VS6. Following closely on thes ZL3RB/Mick became active and due to John's absence, and his own splendid efforts quiet QTH has been the mainstay ever since. We understand ZL1AH/John will reactiva PROPAGATION de Rod Newkirk W9BRD- "The North Atlantic propagation path to Europe highly succeptible to Magno-Ionospheric irregularities which do not significantly effect South American & Pacific Paths according to Termann" Guess this accounts fo those shut-outs while HCs & KH6s BOOM in!!! Thanks a lot Rod - Most interesting!!! SILENT KEY VP7NM/Chas - 160 Meter Hams especially, are saddened by the death of Ch "C.N." Albury. Many of us have his card confirming 160 QSO and are proud of it. He put the BAHAMAS on the 160 DX map. His presence and activity will be greatly misse IN "ZL" LAND EXCITEMENT WAXED HIGH Per ZL3TB/"Mick" and W6KIP/Alex who report that during the CQ/WW/160/cw/Contest, Mick/"There were so many 160 meters signals on bo ends of the band QRMing each other from 0700-1200GMT - I never thought I'd live t hear the 160 meter band dish out signals like it. Although I was the only one to contact W6KIP, - ZL3CO,BC,OX all heard an earful. Stations were too numerous to l Stronger signals were W6KIP,WA6CDR,HC1AGI,W6HXT,W6HRG,K6SHG,W6HG,WØGBV - all audi to 1200 when I fell asleep!! - 5 hours of Dx signals without a break or fade -Wow! ZL3CX Hrd 6KIP,K7HB,W6HG. On Sunday Band was poorer but I hrd WØNWX,2PIQ,W1BB/2 [illegible] would have gladdened your heart to hear that crowd - FUN!

160 DXPEDITION ????? --- K4KSY/Rev. Drayton writes "Wonder if any of the gnag woul
be interested in a DX-pedition to one of our Barruer Islands down off the Coast, W
lived there a couple of years and found conditions absolutely ideal!!! The place i
EDISTO Island - 1½ miles off the lower South Carolina Coast. ---Heard Low Fy Dx
th e that is considered practically impossible elsewhere - European BC stations
consistently - "Scads" of stuff on 160 and 80 -- Next winter possibly some of the
Yankee group going South might like to stop off for a weekend - have access to an
authentic ANTE-BELLUM Mansion just 20 feet from the salt water and poles already
set for antenna. Only power line nearby is the small one to the house. Nearest
neighbor is 2½ miles away. Perfect for anyone DXing in a 50 watt way!!!" ---Anyone
drooling besides W1BB??? Thanks a million Drayton and We'll see!!! (Incidentally
watch for some 160 meter dope in "73" Magazine in near future!!!!)
MISCELLANEOUS TOP BAND NEWS ITEMS -------K7ICW/A1, had a "Ball" in Dx Contest - wk
74 stns/29 sections including HC1AGI & KH6IJ for Total 4500 Points - Hrd XE2OK,
WØAIH/VE3 and says "These 10th wave verticals are for the birds - altho better'n N
NUTIN!!! -------W2UKS/Bill, in that ideal 160 meter QTH on an island, has 2 feet o
seawater all over the place during that March Northeaster -------Steven GIBBS/GM3K
of Ex ZC4AK fame!!! reports from Benghazi, Libya, hearing on Feb 11th, 4 Continent
VE3BQL/SU,ZC4PB,SVØWZ and W1BB/1 plus a lot of other Europeans. Conditions not too
good however/ Steve is 160 meter Dxwise famous for giving two DXers their 160 mete
WAC. W1BB/ZC4XP, March 3 1953 and EXACTLY three years later (As he points out) -
W8GDQ/ZC4AK March 3 1962 -,quite a coincidence - thise dates eh???? Thanks Steve
you prove that it is impossible to keep a good DX man down and we hope to hear you
one of these days as 5A sumpin or other!!! -------Kn3RKD/Ron/-3/10/62 heard W1BB's
1809kc CW 579 on a 3 band citizens USL transceiver converting into a BC radio on
dual conversion!!! Can you figger that one out??? ------- HC1AGI/DON/(W3EIS) lists
CQ/CW/CT QSOs on 2/24: :W2FYT,K2BWR-3RFA,ECR,4KFC,YHD,SQE-5DWB-6KIP,HRG-K6HXT,WA6CI
7UBI,K7ICW-8HGW-ØNWX,CDP,EWH,GJI,SDN-9WJB,YTV-KH6IJ ---2/25: :1BB/1,TX-K2PNF,DOT-
3G[illegible],EIV-[illegible],4AH-K6PJY-8GDQ,GIY-K8HBR/8-9YT-ØGBV-YN1AA-VE3KE-HR3HH-XE1OK, and
Co[illegible]amts Don "Called 10 U.S. guys for each one worked - never heard so many TIN ea
in my life!!! -------PY/SWL/Rolf Rasp Feb 25th hrd 6KIP-K6HXT-4KFC,HYD-K2IUT-8HGW-
K8RRH-9TGB-ØGBV-KH6IJ-HC1AGI- also unidentified VE2 - excellent Rolf-Tnx!! -------
WØAIH/VE3 Rev Paul - Has new antenna ideas for 1962/1963 - hope summer Wx very
conducive to installing same!! QSO'd KH6IJ in CW/CQ/WW/160/DX Contest - also G6BQ
March 4th - CONGRATULATIONS!!!. -------Sven Elfving/Polar Bears Radio Club/SWEDEN
Hrd following Dx this season;;VE3BQL-HR3HH-CX3CB-EP2BK-WØAIH/VP9-VP8CQ-VP5BP-ZC4PB
OK1ACU-UB5WF-HB9QA-VE1ZZ-GI6TK-EI9J-W1BB-5N2GUP-HC1AGI -- VFB Sven - and yout SWL
Bulletin also very nice -Tnx!!! ------- G3PU/ERN CONGRAYULATIONS!!!! His 160 meter
Dx Score at the end of the 1961/1962 season is WAC, 37 Countries!! Greatest
disappointment this year -,almost but not quite QSOing HR3HH. On Feb 18th Ern QSO
VO1FB and Hrd VP5BP ------- THAT 160 Meter ISLAND Dx PARADISE we mentioned last Mo
W2UKS/Bill's Town OCEAN CITY N.J. sure took a pounding in N.E. Storm in March - s
"We've really had it with terrible storm and excessive high water mark - hundreds
of cars a total loss -250 homes damaged beyond repair - dozens of homes washed out
to sea - dozens destroyed by fire -loss of life 5 including C-D worker, Chief of
Police - I have been helping with Emergency work 75 phone and CW. My own car had
$300 water damage - what a mess!!!" ------- Bing/WØCDP reports 1961 CQ/160/CT,,
160x43 - 1962 163x45 including HC1AGI,KH6IJ,XE1OK but gripes!!! "Ws too busy on 1
and NOT covering high end, comments, I am thinking of a strictly 160 meter statio
and location with special low noise receiver" -- Bing pleads for boys to cover TO
END of 160 says "Istill believe I can get signals across to Europe from We if tho
fellows will just listen - this view also shared by W5SOT,DWB,6KIP Etc. I'll be o
1998 each weekend"!!!(See G3PU Report for March 18th) -------W9PNE/BRICE, heard
[illegible] EP2BK/449 during CQ/CW Contest QSOing EI9J - thought he was a pirate so did
c[illegible] him - alas!!! Contest 105x38 VFB Brice. ------- K7HDB/DICK - during CQ/CW/CT
blew H.V. Transformer, of all times!!! Quite interested in ITBY reporting - says
scientists find "Light radiation belt similar to Aurora Borealis" Over Oregon --
believes that this may account for blanketing 160 there - Could Be!!! This ioniz
belt goes apparently around the Globe - wtite K7HDB for further info if intereste
PY/SWL-ROLF RASP -Still listening - no Dx lately -------G3OQT/Dick - is to be
Congratulated on several crossings this season. QSO'd W1BB,UO5AA,1PPN-2[illegible]WD-UB5WF
VP5BP-VO1FB!!!

HAPPY SUMMER-HAPPY HAMMING and 73 de Stew

April 1, 1962

!!! --- 160 METER FLASH de W1BB --- !!!

THE EP2BK story!!!!

EP2BK works VE1ZZ -- W2IU -- W1BB -- WØVEH/VP9 !!!!!!!!!! FLASH!!! On 160!!!

Everyone was thrilled with the appearance of EP2BK/Bob at ABADAN, IRAN - the FIRST time that Country has ever been heard on 160 - QSOing quite a few Europeans

THEN!! during the CQ-160-WW-CW Contest, FEB24, a brief, not too good but definite contact was made with VE1ZZ/John, for the FIRST [redacted] EP/VE -- Congratulations to both Bob and John!!!! During the same morning EP2BK was also heard by W4KFC/Vic, (And others) peaking 588 and thought to be a "Phoney" because of his strength - however, EP2BK also heard W4KFC at 589 and SWL'd

Following this, on March 8th EP2BK arranged regular transmissions and schedules with W1BB (First at 0300, then 0220Z, to attempt a W contact and QSO - finally making definite, but "Touch and Go" contacts with W1BB on March 22,24 and 26th - but no "Solid" 100% QSOs.

FINALLY!!! on March 27th at 0150Z, the FIRST ever Solid 100% QSO, EP/W was made between W2IU/Sam (Ex W9NH) and EP2BK, and followed by W1BB as No.2 at 0220Z their regular schedule time. Bob's signals varied greatly from "nil" to 449 & 53 peaks. W's maximums were W2IU/339-W1BB/449.

IN ADDITION -- Bob/EP2BK also contacted WØVEH/VP9 at 0245 for a not too solid nevertheless definite QSO!!! WØVEH/VP9 peaked at 569 in IRAN!!!

CONGRATULATIONS !!!! Go to Bob/EP2BK for his splendid operating, patience, and perseverance!!!! And also to "Sam"/W2IU for this FIRST!!!

Dick

CONDOLANCES and sincere regrets go to W3GQF/who indirectly made the W2IU QSO possible, when he made this earlier than usual schedule with EP2BK, invited W2IU in on it, and then for some reason or another failed to make contact himself. Also to Charlie/W2EQS, who came in on the opening rather late, and also failed to make itto IRAN. This was all the more disappointing because of the fact that this was EP2BK's last morning on the air in IRAN before closing down and leaving the Country. Also to Shely/W3RGQ who was also heard to make a call this last AM.

BOB accomplished this VFB feat of 160 meter DX using an 813 at 100 watts - into a 15 meter Lazy H antenna 75 Feet high - with an AR88D receiver. He has also worked many Europeans, and we are anticipating a complete report from him soon giving the complete story on his 160 meter acfivities!!!

THE HC1AGI 160 Meter DXPLOIT!!!!!

Another FIRST!!! when HC1AGI and G6BQ QSO'd at 0610z March 11th for the FIRST ever HC/G QSO on 160 Meters. G6BQ was 449 in HC-land. CONGRATULATIONS to both Don/HC1AGI (W3EIS) and Jack/G6BQ VFB indeed!!!

160 sure has been "places" this season!!!!!

EL4A --Ken and Traut Bale expect to be on for a while longer this season. W1BB has sent them crystals for 1827½ and also 1803.0 Kcs Watch for them!!!!!

73 de Stew/W1BB

Stew

EP2BK Mailing address now
Box 502 - Springfield, Missouri.
Robert M. Snyder

OCTOBER 15 1962 W1BB - 160 METER DX - Bulletin 1962/1963 No. 1

de W1BB/General Remarks ----- Hello Fellows!!!! - W1BB checking in with you on wha promises to be our BEST 160 meter Dx season in a long time!!! -"Chatter" collected since spring ended the 1961/1962 season will be found herein. The band was more op this summer than last, with VP8-ZL3-VK3-EI9-G6-W6-XE-FP-GW making real news!! FIRS TRANSATLANTIC QSO (W3RGC/G6BQ) was a month earlier than usual this year. Much new with some suprises is expected!! Many stations W/VE-Dx have used the summer to g advantage for new antenna and ground system construction. As the Sunspot Number go lower and lower, 160 Meter conditions get better and better, so we are apparently for an extremely interesting and exciting season!!! SOOOOOOOOOOOOOO join in on t FUN - get those antennas up high and strong for the winter storms - your SWR 1/1 - check your receivers and transmitters for best performance - mark your calendars & store up a bit of extra rest for those "wee small hours" when 160 is at it's best. HAPPY HAMMING - best of luck and best of Dx HUNTING on 160!!! 73 de Stew/W1BB.

ANNUAL TRANSATLANTIC AND WORLD-WIDE TOP BAND 160 METER DX TESTS ----- Will be held as usual, and this year on Sunday Mornings, December 2 & 16th --January 6th & 20th February 3rd and 17th, from 0500-0730 GMT. During these tests, special efforts will be made to contact European, Asian, African and other 160 meter DXers throughout Transatlantic World, for general DXing, and particularly to also make observations and to contribute to the study of propogation conditions on these frequencies. The ANNUAL 1.8 Mc Activity since 1932 has been planned as TESTS and not Contests - as an unusual operating activity, without competition. W/VE stations will call "CQ DX TEST" the first 5 minutes of each hour, and each alternate 5 minute thereafter, LISTENING the second 5 minute period, and each alternate 5 minute period after that unless working or hearing DX. Dx will of course, use the opposite periods for transmitting and receiving. When DX is coming thru, the idea is to work it regardless of time periods.

Most W/VEs East of the Mississippie River will operate 1800-1825kc segment, while those WEST of the River will use 1976-2000kc, clustering near 2000kc. Most DX wil usually be found bentween 1795-1850kc, while some will be found near 2000kc (ZL). DX stations should cover 1975-2000kc carefully for unusual and choice W/VE Contac W/ stations should send their reports for these TESTS to W1BB, for tabulation, Recording in the booklet " THE 160 METER DX HISTORY" in the making, and especiall for transmittal through BULLETINS to ARRL-CQ-SWM-Western Radio-DX-RADIO ZS and ot G,GM,GI,GW etc should report to L.H. Thomas, DX Editor Short Wave Magazine, 186 Winchelsea Rd., Hastings Sussex, England. Other DX may send their reports to eith of the above. Please DO report accomplishments, Happenings, Rumors and Data!!! W1 will award Certificates to all DX stations working W1BB, if requested, and to DX SWLs who send 7 verified reports of W1BB's signals - also IF requested. Certifica for past work on this band are available - contact W1BB for further details. ---- Working DX on 160 is indeed challenging, and extremely interesting. Conditions su as static, BC Harmonics, QRN, QRM, LORAN, NOISE, QSB, all require extreme patienc with perseverance, a top-notch station and A-1 operating ability. There are many FIRSTS yet to be made and many unusual conditions to cope with. One gets anew the thrill of early Amateur Pioneer days, when HAMS were relegated to "TWO HUNDRED METERS DOWN", and Paul Godley made the FURST Transatlantic Crossing via Amateur Radio, thereby inaugurating the first of what has now become Yearly TESTS, enjoye by many 160 meter enthusiasts. In addition you will make many friends, among the finest and most sincerely devoted HAMS you would ever care to meet!!! OK?? b.c.n.

160 mete

CONTESTS ----- W1BB and many other W/VE 160 meter HAMS will give special/attentio to DX stations during the ARRL and CQ DX Contests and also the special CQ 160 Met CW Contest, (Which will come earlier this year - watch CQ). W1BB invites any special schedule arrangements from any of the DX gang desiring to make a real try for an extra band multiplier. AIR/MAIL your intentions to W1BB in this respect, i advance if possible, otherwise W1BB will generally cover all 160 DX Frequenc

W1BB 160 METER DX BULLETIN -----Will be issued during the 160 Meter DX Season, as material is obtained, to the above/mentioned HAM magazines, via air-mail, so that they can publish whatever may be of general interest & the most important informa on TOP BAND activities. There is no mailing list to individuals, although W1BB wi oblige and send copies to any W/VE stations supplying s.a.s.e. - and to any DX station who reports regularly to W1BB on 160 Meter DX Activities - also SWLs. W1B also will send FLASH news items to Magazines on any special events and happenings

FIRST 1962/1963 TRANSATLANTIC QSOs ----- Congratulations to W3GQF/Dick/359 who on Sep. 16th QSOd Jack/G6QB/449 at 0524z one of the earliest openings on record!!! Second crossing was made by W1BB/1/489/0405 from farm in Maine Sept. 29th Third was K3MBF same date. Of interest FIRST SIGNALS were heard "Across th Pond" by G/SWL/Norman Smith Sep 1 0423 W2EQS and W3GQF/459 - Sep. 2 W3FBV/weak CQ answered by G3PU and G3OLI - Sep7 0353 W2IU/459 - Sep8 W3GQF-Sep9 G3PU Calls W3FB Looks like an excellent season shaping up with openings so early!!!!

SUMMER DX - VP8GQ!!! ----- 160 Meter DXers were thrilled with the lowering SunSpo numbers and the pick-up in 160 Dx to hear VP8GQ in the South Orkney Islands comin thru frequently this summer QSOing a number of W/VEs also EI9J "Paddy" to whom go special congratulations. Credit for VP8GQ's appearance and following info goes to Dick/W3GQF who when working him on 80 June 30 asked him QSY to 160 and try - the was "right" and results immediate!!! W3GQF/589 VP8GQ/579 0300z/1827k for a solid 20 minute QSO - contacts followed with W3RFA & W2FYT. W3HKE almost made it but no quite as signals faded. July 1 W3GQF again and W2KQT -July 2 0300z W2IU,KQT,TKG, VE3GP,DU-July 3 EI9J/0202,W1BB,W2KQT,VE3QU and W3GQF for his third QSO with VP8GQ W3GQF valiantly but unsuccessfully tried to hook VP8GQ and W6KIP/Alex but he was out of town. Finally July 9th VE1ZZ/VP8GQ QSO. Signal reports averaged S78 for W/ and 548 VP -- these results still further confirm W1BB's often expressed views th the band is best open to the South summers, after having had QSOs with CP5EQ and other South Americans during prevoius summers. These latest W/VE/EI QSOs confirm!

Further confirmation comes from PY/SWL/Rolf Rasp in R.J. Brazil S.A. who reported JUN 16 0620 hearing W6KIP/Alex/579 QSO W6ZH/Herb/559-589 1999k. "Their signals we so clear and loud that I did not miss a word. Never thought 160 signals could be loud"!!! To which could be added that W3GQF/Dick QSO ZL3RB/Mick MAY6 0955 - VFB!!

VK 160-Dx!!!!! ----- The VKs with their new liberties on 160 are pushing for 160 Q with W/Ve & Dx. On Sun AM SEP2 W2FYT was suprised to hear a weak stn clg on his f A sort of contact was established before QSB out. The station turned out to be a call uncertain - believed to be VK3AMZ on 1803k at 0614 - since VK3AGS has been i identified and it is believed he was the station. News spread and the 160 boys we "HOT" after this exotic DX!! On Sep9 VE3DU W1EFN-Maybe others- hrd ZL3RB 1015z on CW & A3 but no QSO. Then the band to VK Cracked open!!! On SEP20 K3MBP W1EFN W2FY broke the ice for FIRSTS QSOing VK3HG. On FRI AM SEP21 W1BB/1 on farm in Maine an W3GQF QSOd VKeAKR whose signals peaked 579 staying in from 1000-1100z Steadily!!! W3SO/459 QSOd VKeAKR/359 SEP28 as did Chas/W2EQS for his 34th Country!!More to co

ENVIABLE RECORD/W3GQF!!! ----- Dick kept the 160 band warm QSOing this summer & F VP8GQ-WØVEH/VP9-FP8CA-ZL3RB-XE1OK-G6BW-VK3AKR-G3OOT-GW8PG-G3ERN-G3PU and has had least one W6QSO every month!!! Says maybe odx will open up DX year/round. Suggest helpful "Markers" - DHF/1831k- and when WWV is off::: CMA/2500k/Prague/Czech/1000 1000cps/A2-or FFH/Paris/300w-HBN/Swiss/300w -JJY/Tokyo/2000w-LOL/Buenos Airi 2000w-MSF/Rugby/500w-ZLFS/NZ/30w - Thanks Dick - nice work!!!!

THANKS TO CHARLIE/W2EQS ----- who is banging away at his key on 20 and typewriter "Airmail", stirring up 160 DXers in foreign lands for this years hunting!! Report PY2BZD-HK1WW-HK4DP,1HV-GC2CNS-OA5G-VP3AD-YN1AA-YU1GM,AD-ZB1BJ-OE3SE-ZE8JJ-HI8XAG all may be on. Some have worked 160 before but others are new ones. Ken/EL4A is hoping to be on 160/but not for sure/expecting new Jr/op etc. - Charlie/EQS sport a new 160 doublet which netted him the VK and intends to do his share of Dx Hunti

EARLIER THIS YEAR/VP8GQ/Peter Hobbs (G3LET) ----- opened up one of the rarest 160 spots on the map appearing JAN28 QSOing in order HC1AGI/569/0600-WØIFH/469/0627- W1BB/1/469/0640-WØNWX/579/0653 - FEB17 WØNCS/569/0625-W3FBV/569/0640. Peter is to thanked a million times for these FIRST SOUTH ORKNEYS 160 QSOs. G3PAG, His very meticulous QSL Mgr is to be thanked too for FB QSLing - Peter's address is FALKLA ISLANDS DEPENDENCIES SURVEYS Base "H" Signed Island, South Orkneys - but QSL via

SAINT PIERRE/FP8BX/W1RAN/"Ned"-----Who made 160 so interesting for many of us dur contests says "Grand time on TOP-BAND . Golly could I hear Europeans!! They were S789 and standing out over Ws as early as 0130Z!! VFB NED - Nw we All like try FP

EI9J and 160 in EI Land -----"Paddy" says until recently licenses for 160 were no generally granted, but now can be had for the asking. At request of W4KFC, last season Paddy whipped up quick final/ got on with well known FB results. EI9J wks CW mostly but found 160 refreshing after the commercial and HAM QRM on 80. During CQ Contest "Paddy" was disturbed by many Gs calling him when he wanted Ws & other rarer DX & Multipliers!! "EP2BK answered 2 of my CQs/was so loud/ took him f phony & didnt ans.-ALAS!! Later wkd him however!! Am sure that his signals were t most remarkable of the test!! Now that I have the gear and permit an looking forw to being up on 160 the 1962/1963 season!!!" We'll all be happy to hear you Paddy!

COINCIDENCE ZCs QSOs w/Ws !!! ----- Stephen Gibbs/GM3MBS/ZC4AK now 5A3CJ/Libya refers to p4 W1BB DX BULL No.5 1961/1962 pointing out that ZC4XP was Sid Parks,,, making this comment "Stew-I bet if you examine your log you will find that ZC4AK/ W8GDQ QSO MAR3'62 and ZC4XK/W1BB QSO MAR3 '53, exactly 8 years apart to the day, would be at about the same time GMT" ----- And SURE ENOUGH!!! ZC4XK wkd BB 0400z ZC4AK wkd Willie/W8GDQ 0417z - Pretty close eh???? You boys who want to QSO ZC n know the right magic date and TIME!!! Thanks for interesting observation Steve!!

W1BB----- Celebrating his 50th year of HAM Radio - made his 60th 160 Country QSO working VP8GQ - The next 40 to DXCC/160 will be the hardest - although we MAY mak it before "Silent-Key" time!!! Great game this 160 - enjoyed every minute of it

MISCELLANEOUS TOP-BAND NEWS ITEMS -(Dating back to April 1962)(And up to date) 4.
Gswl/Allen/APR8 Hrd Wa2KWH,5SOT,ØCDP,2IU,1AW. George was real excited SWLing ØCDP
His furthest WEST yet!! Still trying for W6&7 -----W2IU/Sam has FB 125' antenna To
QSO'd EP2BK w/1t & should be ready for more FB DX to come!! ----- WØCDP/King/1998k
APR8 hrd G3ERN/129-QSOd ZL3RB/Mick 0820 339/both VFB/Congratulations!!!! -----
K1[illegible]/Paul hd rig trble/missed most Dxtests-got real thrill however hearing FIRST/
JAN7 ----- W2GHK/Stuart(Hammarlund Co) says thru his influence their new HX50 SSB
has 160 coils available!! Expects be active 1962/1963 w/inverted"L" - Best of Luck
-----WØPSF/Larry, Hoisington Kansas hopes activate 1962/1963 after stretch/army. M
hve real FB ant on 400' tower!!! bonu Larry!! ----- GARY/K1KSH, fm USAF/Turner/AF
says 50/50 chance he may become rare 160 DX within 6 Mos-were hoping Gary!! -----
W8HGW/Les,(Parkview Electric Co), who started 160 1961/62-looking fwd to more act
this season/listen fr him!! -----Fm G3ICW/"MIKE" WHITAKER,a fine rpt FEB/10 QSOd
ZC4PB fr his First/Ever QSO on 160!!! FEB/11 QSO VE3BQL/SU -- On CQ/WW/DX/TEST-14
countries-16Mult-9136 Score!!! JAN7&MAR3 each 7 Ws QSOs-FEB/18VE1ZZ&VO1FB - Wishes
Ws would give more data on aerial system& Pwr on their QSLs also wonders why there
are no W/SWLers giving 160 reports to DX boys -- (W1BB wonders too!!!) There just
doesnt seem to be any W/VE 160 meter SWLers!! Also comments "Over the years we hav
noticed a N/S skip with E/W "Cut" - even on 160 - this happens on 80 a lot and is
more noticeable as greater level of activity shows up - not necessarially due to
ATOM bombs - but just normal variation on ionized layers"!! Thanks Mike - fine ide
-----G3LIQ/Denis, gives a list of outstanding stations be rec'd viz, Ws8JIN,NSF,VI
K8HBR-3GQF-K2DGT,IU-1AW,1BB-W6KIP,ARR22!!!!! - HP6AKZ,HC1AGI,HR3HH,E14A,CN8PZ,EP2B
also QSOd/W1BB - VFB Denis and more good hunting 62/63!! ----- G3OJI/Jim,Planning
new 160 ZEPP at new QTH/quiet unobstructed. May try vert kite ant!!! Best/luck Jim
-----W8TSD/Dave, was vy active this summer on the 160 Meter "Mills Trophy Race Net
on Lake Erie. Also participating were K8YOO Chm-K8SCW,W8CUR,K8YON,W8IME,RZ4,OFG,GJ
& K8LFI - VFB boys but how about some 160 DX????? ----- W7UEI/Keith , Expects 2b
active 62/63 w/DX100/SX71.Inverted "L" Marconi- doing lots of W/Coast/SSB - bonu!!
----- W8TSD/Dave, again - says at close of RACES everything went fine and dandy -
al[illegible] there are 500 HAMS in Toledo area and 100 on 160 & More coming - Toledo Emgcy
NET AREC/1812Kc have around 70 sign-in each month's DRILL Sunday 1PM - Excellent w
----- W6HRG/FRANK, reports QSOs between 1.8/2.0 Mc Stns hve occurred each month
during the summer!! Frank QSO'd KØPAU/57-89 at times JUL31 A3 - WA6AJY & WA6CDR QS
him on CW. W3GQF & WØRFT hv been wkd by a number of W/Coast stns during summer! Fr
believes more results would occurr if stns tried summertime DX more!! ----------
SORRY TO REPORT VK4/SWL/Charlie Thorpe, 81 Dawson - Rockhampton-Queensland-Austral
had back operation - wl B in cast fr 6 Mos from HIPS to under arms - must stand mo
making it hard for SWLing - however will neverthaless cover 160 as well as possibl
62/63 season -Great Spirit Charlie!!! Tnx ------- 8/28/62-K1KSH/Gary says "ALHOA"
Hawaii and hopes to get on 160/MARS stn soon - still expecting to "Pop-up" on 160
fm some rare & EXOTIC DX spot one of these days!!! ----- DY/SWL/ROLF RASP,Hrd W6ZB
559/589 & W6KIP/579 JUN/16 and copied entire QSO ----- W1AUQ/"ED"-is showing inter
in 160 & may be on!! FB/Ed -----WILLIE/W8GDQ, checks in 62/63 - hrd weak ZL call b
Tried VK3AKR who QSO'd W8HGW.Willie has 28 Countries- sure nice to have you Willie
one of our old reliables on 160 checking in again for the 160,season!! -------G/SW
Norman Smith re orte hearing W1PPN,2E4S,FYT,IU,KHT,K2PNF,3GQF,8HGW SEP30 ---------
G/SWL/ANTHONY M QUEST/LEEDS/ISWL/G5967 also hrd same stations - except 8HGW-Tnx/To
----- WILLIE/W8GDQ, QSO'd VP8GQ SEP23 ------ FIRST VK ever QSO expected to be
credited to K3MBF - Nice work DAWSON!!! SEP20/QSO/VK3HG!!!! ------- W1EFN/Walt,
sporting special new 160 ANT280' Horiz & 570' wire in vert section 90' off ground
and HT9 190w was SECOND VK QSO on record w/VK3HG/SEP30 ------- Comments also that
W6s come in frequently around 6.30AM LOcal time - sunrise about - with suprising
strengths at times. QSO'd WØVEH/VP9 JUL21 FP8CB/JUL27 - rag chews more than DX but
enjoys DX when possible - worked lots of Mobiles on 160 this summer mostly W8s --
----- GW8PG/Gus, reports operation from G8PG/AmGreasby SEP23 hrd W3GQF, W1BB/1 -
SEP30 as GW8PG "Boy what a party" - Wkd W3GQF and Hrd W2FYT/339-2E4S/569-2GGL/449-
K2PNF/559-2IU/569-2KHT/559 -- Gs QSOing Ws were G3PU,ERN,OQT & 6BW ---OCTY Band
in bad shape - NO DX!!! Gs3LIQ,LLE,PGN.NEO,LYW,OQT,RBP Participating.

FIRST VK/W QSO CREDIT !!!!! Does go to K3MBF - this has been confirmed by W1EFN
The station was VK3HG SEP20/62 o515EST - Congratulations DAWSON!!!!!!

HAPPY HAMMING Fellows - and best of DX on 160 -- 73 de Stew

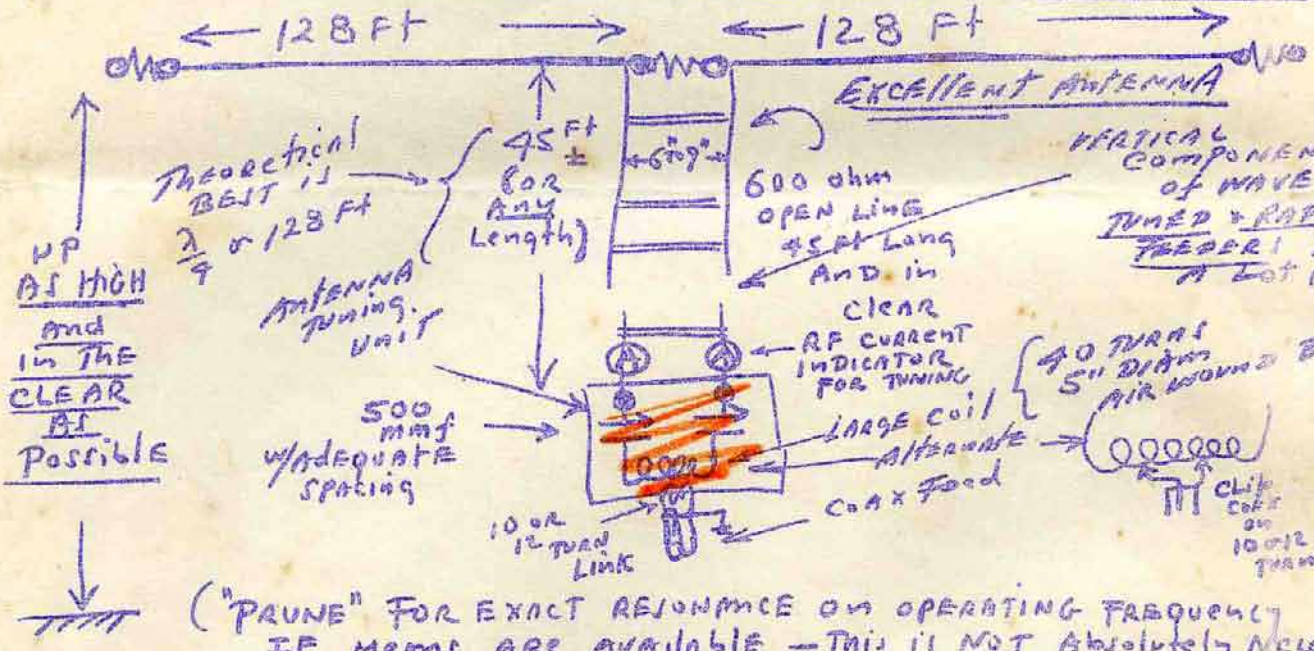
W1BB

ANTENNAS FOR ONE SIXTY

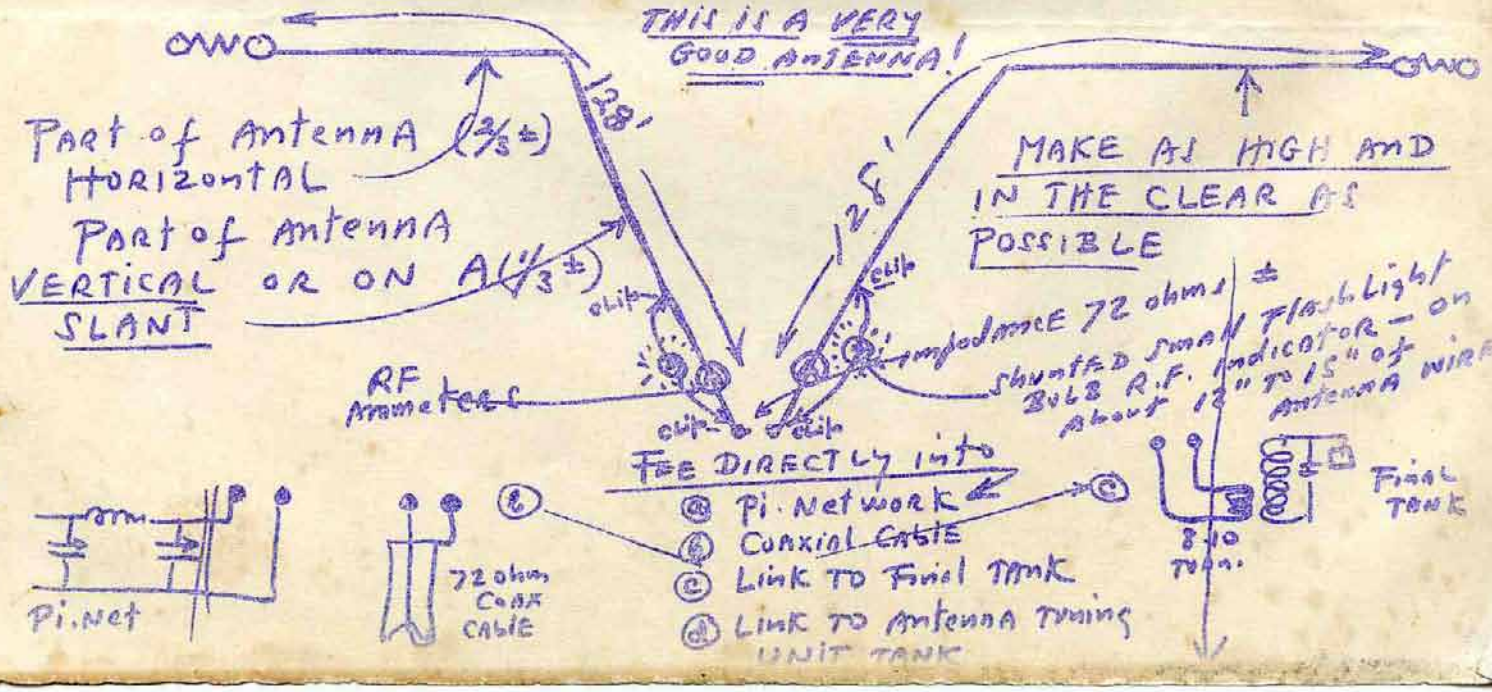
LISTED IN ORDER OF DESIREABILITY FOR BEST DX WORK

NO. 1 — HALF WAVE DOUBLET — W/OPEN WIRE FEED

THIS BEST Antenna FOUND BY ACTUAL TEST!



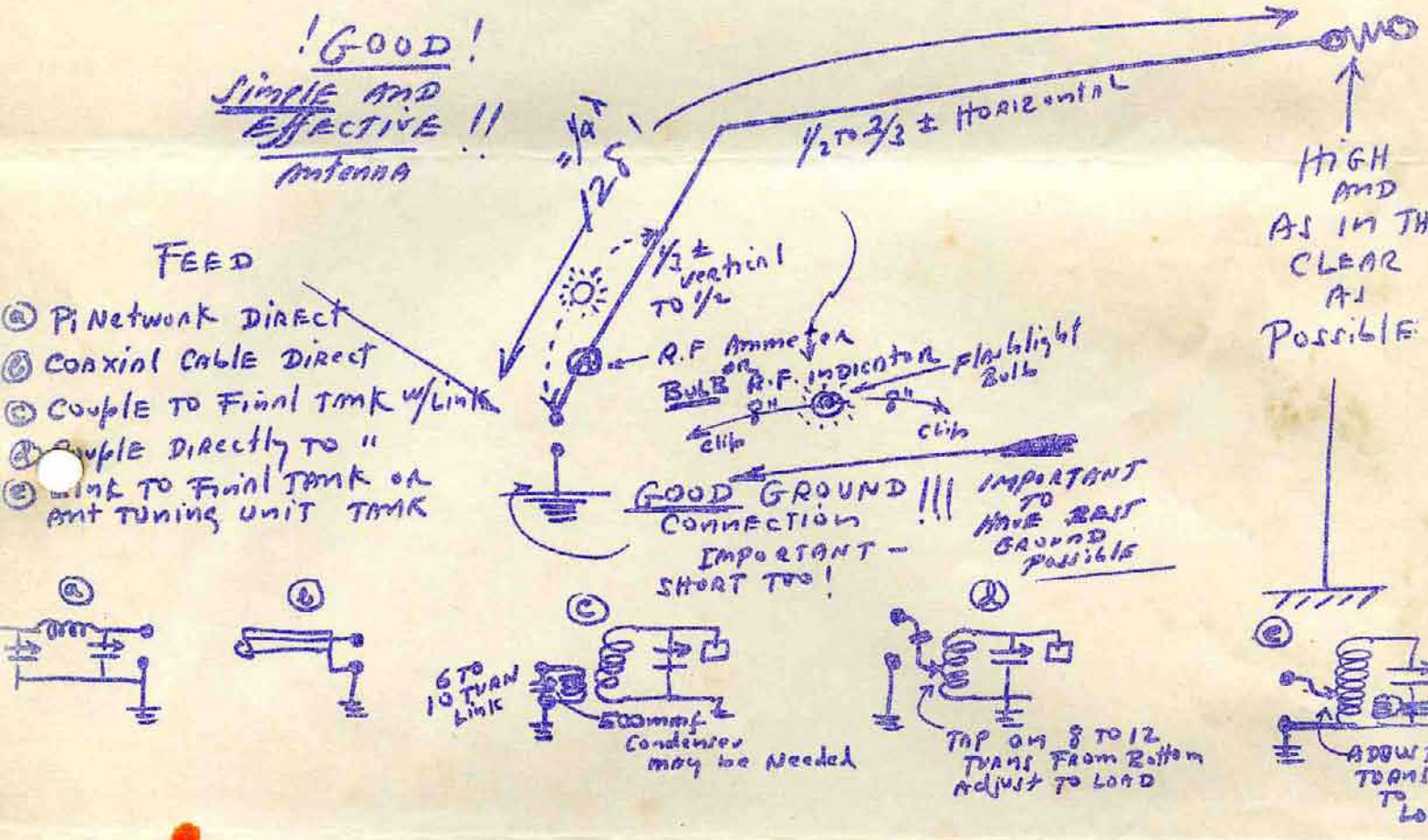
NO. 2 — HALF WAVE DOUBLET — BENT and BROUGHT DIRECTLY INTO SHACK FOR FEED — 2nd BEST ANTENNA



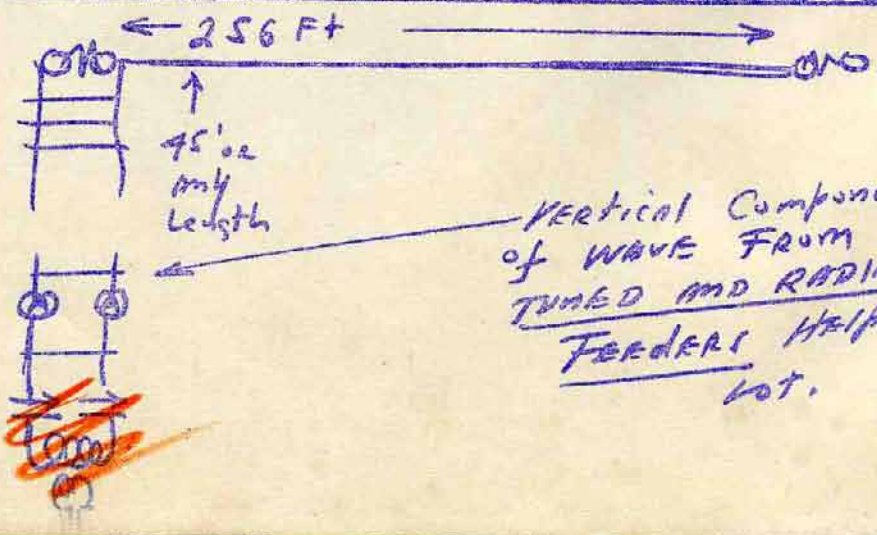
W1BB - ANTENNAS FOR 160 — continued

NO. 4 1/4 WAVE - SINGLE WIRE ANTENNA WITH "GOOD" GROUND 3RD BEST

(THIS IS EQUIVALENT TO Half of Antenna #2)



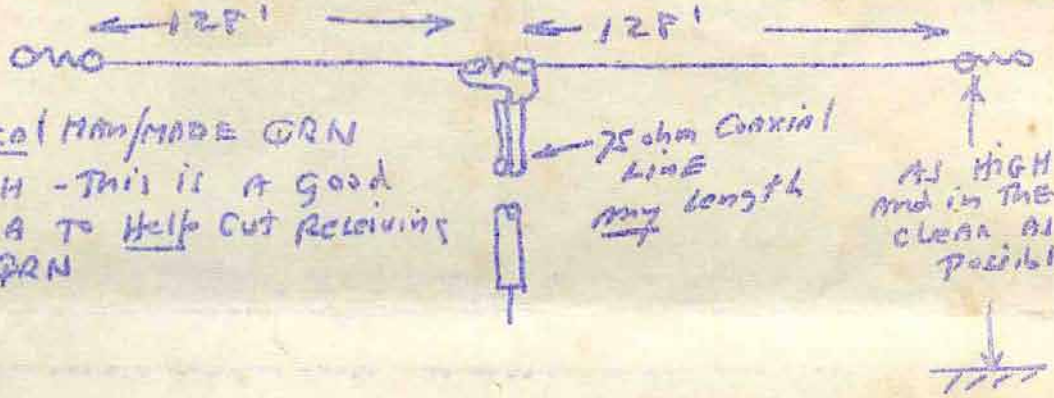
NO. 1 1/2 WAVE ZEPP ANTENNA - 4TH BEST



SEE NO. 1
EXACTLY SAME REMARKS APPLY
AND SAME Component VALUES

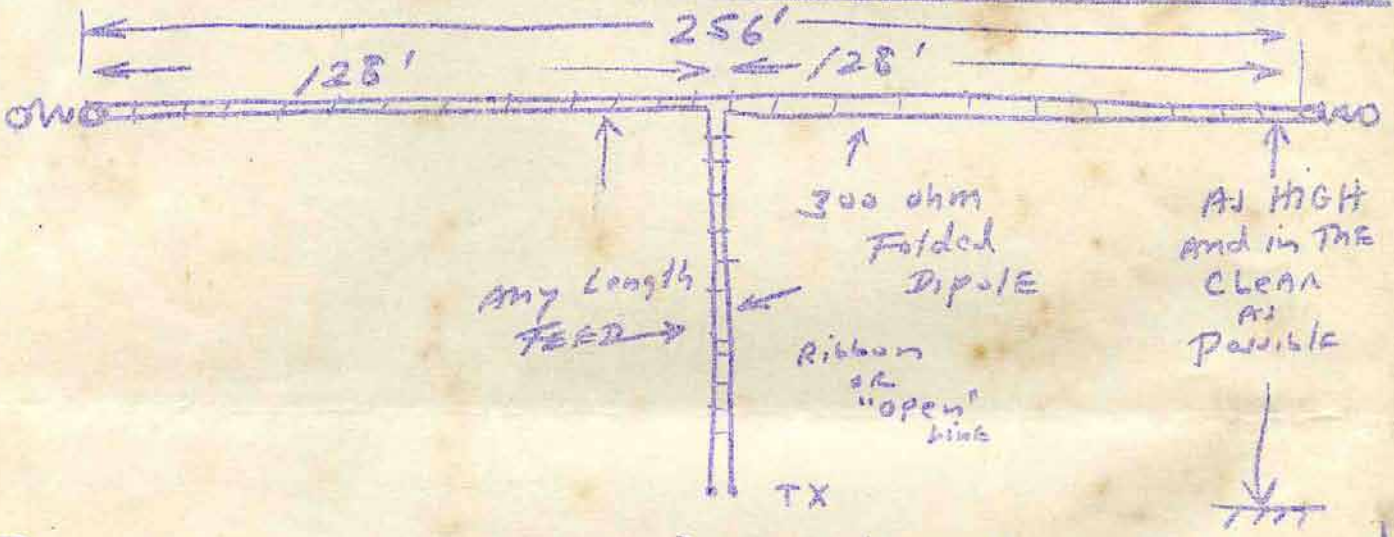
W1BB — ANTENNAS FOR 160λ — (Continued)

NO. 5 HALF WAVE DOUBLET — COAXIAL FED — 5th

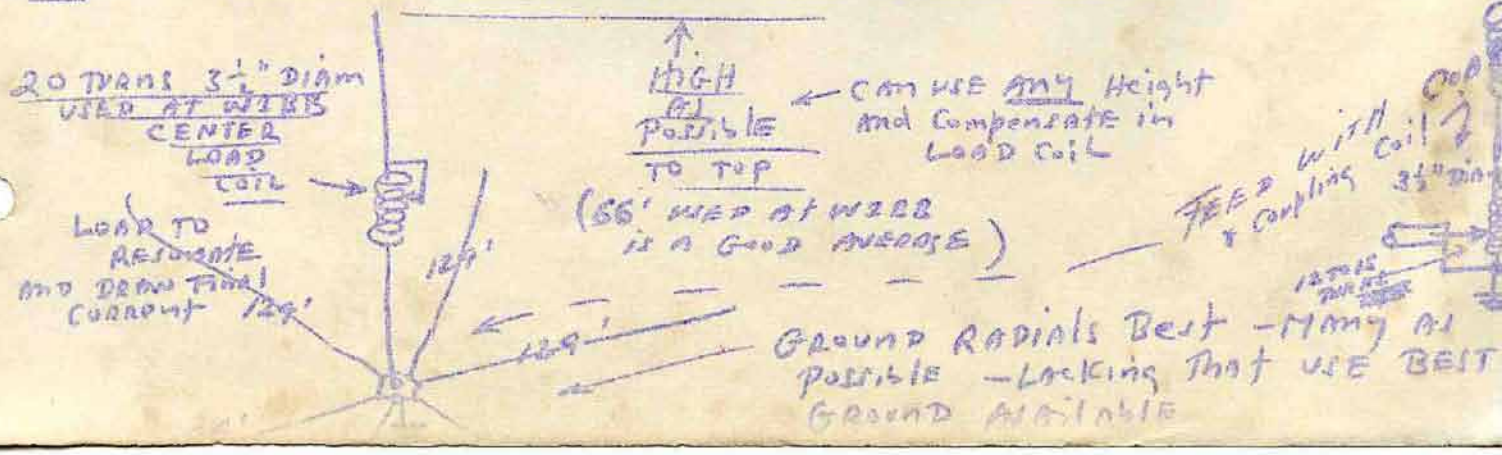


NO. 6 160λ Folded Dipole w/ 300 ohm Ribbon

OR MUCH BETTER!! w/ OPEN 300 ohm Line



NO. 7 VERTICAL 160λ Antenna



December 28 1962

W1BB
STEWART S. PERRY
36 Pleasant Street
Winthrop, Mass.

PYswl/Rolf

Here is all the latest on 160 O.M.!!!

Nothing on 160 this morning - see log - sorry

Starting New Year will make Wed and Fri at 0600

And as soon as OK with you

Make it Tue and Thursday at 0600 instead Wed/Fri - OK????

CONGRATULATIONS yourfine work there on 160

K9GZK also reported to me your hearing him - VFB!!!

Am waiting to hear from you re crystals - tolerances

SOOOOOoooooo again Rolf - HAPPY NEW YEAR

And here's to a contact REAL SOON NOW!!!!

73 de Stew/W1BB

73! de W1BB/Stew

DECEMBER 31 1962 **W1BB - 160 METER DX - Bulletin** **1962/1963 No. 2**

de W1BB/General Remarks ----- Follows, if things have been a bit late, or some of your reports and letters not acknowledged or answered as promptly as usual - Pleas Excuse - Alice/W1DQF, and myself were called home from 12 days vacation in KP4/KV4 lands, when very recently, Dec7th, we had the second death in our immediate family with the passing of our Son's wife, 30, leaving two little children, which we are now helping to care for - all of this upset us and our normal activities in many ways - however, we expect to be back "on the beam" again real soon - thanks for yo patience and understanding!!! DX this fall was good up to October 1st, with the earliest Transatlantic crossing on record. Then, rather suddenly from Oct1 to Nov25 Dx was Nil - since there has been Dx but spotty - no real good openings only a couple "fair" ones - Why conditions which started out so promising, have be so much poorer than expected, no one knows - some feel the Pacific Nuclear and USS tests might have been responsible - others that it is just the "Way the ball bounc , normal unpredictableness of 1.8 - which incidentally makes it interesting and al intriguing. The First scheduled TRANSATLANTIC DX TEST Dec2 washed out - the 2nd De 16th only fair although participants were not lacking on either side!!! - At any r the frequency of openings lately has improved a wee bit, and lets hope for the bes Dont forget the CQ/WW Contest on 160 Jan26/27 and best of luck!!! This should a grand event!!! SOOOOOOoooooo Best wishes to all for a HAPPY NEW YEAR, with l of HAPPY HAMMING - especially if it is on 160!!!!!

W1BB/Summary LOG of 160 Meter DX Activities!!!!

Oct21 --- Poor morning to Europe-fair to South - medium w/ve activity - Wx- VFB/warm/clear/calm -felt like gang just "warming-up" for the real Dx ahead!! Alth VP8GQ put on a fine performance!!! W/Ve p/W1BB,ME,2EOS,UWD,IU,3GQF,CBM,FBV,40M SGRQ,K8IQQ,MRB,9FNE,ØCDP,RFI,GDH,VE2UQ,3QU ---p/ etc stands for Participants!!!)

EAST:::: All-not a sigNAL of anykind and a pox Gus/GW8PG decided it was time to degenerated into "Rag-Chewing". p/G3IGW,OUK/A,OIT,MYI,REP,RAA/A,OZF,OQT,AGG-6BQ-MF,PG/A. Tnx Gus! SOUTH:::: VP8GQ 1800.5 our "Southern (Dx) Star" 0500-0630z weak QSB zero/459, av 339 QSO'd Ws,1ME,TX-2EQS,UWD-ØCDP-9PNE(?),1BB-K8MBR in order was hrd by Py/Sw1/Rolf Rasp!! Says "Hrd VP8GQ QSOing one W after another, BUT the whole time through I did not hear one TRACE of the Ws - only VP8GQ- dont you think this propogation was rather strange??" Yes Sure Do Rolf!!! For Roy/1TX his QSO with VP8GQ was his 10th Country!! (Oct.7 VK3AKR his 9th!) (p/ means "Participant

OCT28:::: Extremely poor Dx Cdx. WWV sending "W3" Wx,clear, cold, windy,PB.-W/VE:::: Ws,1BB,ME,TX-2UWD,KHT-3GQF-8HCR-9PNE,CUY-VE3DU participated, calling and listening for Dx but nary a peep and by 0600 most had QRT!!

NOV4:::: Again Cdx Vy poor-no Dx at all except W3GQF reported hrg DL1FF 0530 559max - HI8XAG & 5B4PB expected but no show - There seems to be some barrier, that signals of any strength do not penetrate, since Oct1 - Nuclear Tests?? W/Ve:::: p/Ws,1BB,BHQ,ME,WY-K1LTM-2EQS,IU,KHT,UWD-Wa2IUQ-3GQF-4CXW-Wa6CDR-8CIM-K8RYU-9UKV,PNWX,VXO-VE3FF EAST:::: No QSOs but G8PG/a hrd W1BB 229-339 o530z. DJ1BZ, GM3PRA

NOV11:::: Condx still generally Vy poor especially to East. W/Ve Participan dropping a bit also. W/Ve:::: p/Ws,1BB,UZ,TX-K1LMO-2FYT,KHW,JGQ-3GQF,FBV-K7DEK/BIQS,GDQ-K8WGW,BBI,IQQ,HSW-9PNE-K9JDK-ØIFH,GBV-VE3DU,QU. EAST:::: Cdx vy poor w/ no European markers or Dx coming thru!! However GW8PG reported hrg W2FYT briefly a 0438z clg CQ DX -- Gsw1/QUEST hrd a N. African but no other Dx. p/G3IGW,HR3HH,GM3PA EI9J,DL1FF,VP8GQ SOUTH:::: VP8GQ came thru weakly 0440z QSOing W2FYT. Our old friend HR3HH/"Hal" put on best show of evening 0600-0700z QSOing Ws1BB-8HGW-K8IQQ-ØIFH,VXO and others. WEST:::: K7DEK/7 provided some welcome Western DX and new states for a few QSOing VE3QU,DU-W8GDQ,HGW-K8IQQ and others - Was hrd by W1BB S2/3 Wa6CDR active QSOing WØGBV/others!!

NOV18:::: Cdx still generally poor-QRN rather high-No Dx whole evening exce VP7NY "Local-Dx" W/VE:::: p/Ws 1BB,2DHQ-3FBV-K3MBF-4IQU-8TSD-k8RYU-9PNE-VE2UQ,3M 2FYT,UWD,GGL,EQS-1BHQ. EAST:::: Gw8PG/Gus says morning a "Dead-Loss"-Blizzard cu power lines-became good cook over fireplace but no Dx!!! SOUTH:::: VP7NY/Lowell put on a fine show wkg Ws 1BHQ-2FYT,UWD,GGL,EQS-3PAP,GQF,PMV-8QWI-K8RYU,MRB-9PNE-VE3QU,AGK. Cdx so poor his sigs varied 000-569 quick QSB 2 hours much of time unreadable in QRN/QSB WEST:::: A fine clean out FIRST!!!!! when the OM/XYL team WB6AFH/WB6AFI (Ex W9VER/K5VLG) QSO'd VR30 Christmas Island on 1991k Nov18. VR30/44 6s/559.-----(Under call K5VLH QSO'd ZL3RB&KH6IJ Fm N. Mexico early 1962 and as W8FQ QSO'd G-VE-VP7-KP4-KV4-YN4-YV5 in past years) CONGRATULATIONS!!!!!! Jean and Dean! Understand VR30 will be on regularly when he gets new Tx now being built by W6VHY. Note!! W6HRG also worked VR30-Details not rec'd hr.

Nov25:::: AT LAST the band showed some signs of opening up and quite a numb of Dx QSOs resulted - our best morning in some time!! Willie/W8GDQ says dont forge to cover 1876k fr ZLs and also claims "MOONBOUNCE" QSO w/W1BU on 160!!! Hi!!! W/Ve:::: Sam/W1BU says QSO w/W8GDQ was 1st 160 QSO since coming to Mass fm Ohio 1 °40s - using pp814s/rig fm 20yrs garage storage-has tilted ant 200° high NE end-ni to have Sam on again!!! Maybe sometime Sam and Willie will explain this "Moonboun stuff on 160!!!!! p/Ws1BB,ME-2FYT,EQS,UKS,IU,GGL-3GQF-6KIP-Wa6CDR-8FW,HGW,GDQ-ØVX(VE7AN. EAST:::: G3IGW hrd W2FYT,GGL,IU and briefly contacted W1BB. UO5AA on 160 G8PG/Gus Qsb/Qrm heavy sigs weak but wkd W1BB 1st time & Hrd W2IU -G3LHQ hrd W2FYT G/Sw1/QUEST h/1BB449,2GGL349,FYT339 - nw hrd 19 Countries 160-logged 15 Dls this yr ---Gsw1/Smith h/1BB,ME,2IU,FYT,GGL after some blank weeks!! ---- G3OQT/D "The Dx Blanket which covered everything OCT & NOV seems to hv lifted nw-hrd W2GG 449,2FYT/449 & 1BB/549-Qrm Hr Fantastic!!! Making himself an 80° vertical Beer Ca ant wl top load it"also wonders why there are no W Sw1s- says Gs wld like rpts!!! Gsw1/Iball h/1BB,ME "Have never known such activity/QRM over here!! p/G3OQT,PU,IG PXZ,LIQ/aMYI,REP,JFY,OZF,NEO-6BQ-DL1FF,DJ1JZ,UB5CI,UB5WF,UO5AA,OH2NB,OK1KAU,KSO,AI ZC YD,KVE,2KJU,5B4PB,OY7ML SOUTH:::: W8GDQ/Willie wkd YV5AGD as did W1BB,W3GQF and others. VP2VL/WØNWX Bob was on QSOing 1BB,3GQF,8GDQ and many others - believe Bob intends retire VP2land someday and work DX to his hearts content alongside next door neighbor KV4AA/Dick who sure sets a good example - Dick an ex/160 Dxer is washed our on 160 by "Regs" while on VP2 Island few miles away VP2 can wk 160! W2UKS/Bill rpts wkg VP2GX/1810k 2334z-VFB!! WEST:::: W6KIP/Alex had 17 eastern "Without trying" hrd HC1DC-VP2VL/S3 -Couldnt find YV5AGD" - Sri Alex!! Next time.

Dec2:::: W1BB was in KP4 land/unable observe conditions but understand were generally poor!! W2UKS/Bill rptd band vy NOISY & "Dx sure is scarce to date"!!! has HQ145x new/March '62 for sale $175.00!!! VFB Cdx. EAST:::: GW8PG/(G8PG)Gus 1MB29-has driven 300 miles this winter to try wk 160 fm GW8PG --- Gsw1/QUEST says "N Dx Hrd" p/G5IGW,OLI,OQT,MYI,GRL,OLB,LIQ/a,PQA-6BQ-GW8PG-GM3ITN,PBA

Dec9:::: W1BB returning fm KP4-KV4 lands unable ck conditions - however fm rpts apparently not too bad!!! NOTE- the night before DEC8 W3GQF worked an O.T. an famous early 160 meter DXer - and noted DXer all bands G2PL Peter - and hrd G6BQ, 30QT,IGW-G6GM and DL1FF!!! W/Ve:::: W9PNE hrd DL1FF259 15 mins-G3PU339 Briefly - EAST:::: Gsw1/Allen W1ME,TX-2EQS,UWD,FYT,GGL,KHT,IU-VFB!! W1TX wkd G6BQ-G3PU ---E G8PG/a/Gus hrd 2FYT449-2IU229,EQS559,GGL339,UWD339 sigs weak/QSB Predawn Peak0612- postdawn peak0747 ---"Harry"/G6GM 78 Yrs yound putting out VFB Sig as usual!!! --- Gsw1/QUEST; Ws1ME,TX-2EQS,FYT,GGL,KHT,UWD says PAØPN,SS-OH2YV,NB ou!! ---- Gsw1/S Smith Ws1ME-2IU,FYT,GGL says Ws 3 deep between 1806-8 & Fish- fones real bad!! ---Gsw1/Iball-1TX359-2FYT569-GGL559-EQS449-IU459-3GQF339 using HC1155 Rx. SOUTH:::: VP8GQ trying QSO Gs DLs/N.D. --- VP3AD/Dad hrd W2FYT/ 569/0410-UWD449/0425-UKS459.0436-8IRE559/0432 - Called all of them on 1801/N.D. Dx/Participants G2DCG-30QT,PQA,PU,LIQ/a,PHO,OIT,CHN,REP,IMV,PIQ,ERN,GRL,NHE,PC KFT,IGW,NJR-5JU-6GM,BQ,QB-8PG/a,JR-DL1FF,FL-PAØPN,SS-OH2NB-GW3PPF,VP3AD.

Dec16:::: This Second Annual TRANSATLANTIC Dx TEST was much better - but still spotty and with lots of snow static & Loran - Nevertheless quite a few QSOs ensued between 0500-0800 with a fwir number of participants and several outstandi QSOs. As a "Warm-up" W1TX on the 14th QSO'd G2PL,3ERN,6BQ,6TK and 15th G6BQ,3ERN, PQA W/VE:::: Roy/1TX wkdG3PU -has new VFB antanna on only 50' Lot - results are what count!! W9PNE/Brice hrd DL1FF 30 min peak 239 also G3PU 0700z 339 rapid QSB. NOTE!!! Dec 15 Brice wkd VP8GQ solid got 570 gave 459 for his 35th Country!!!!! (Coincidence - VP8GQ was also W2EQS/Chas's Nr 35 Country!!) --- CONGRATULATIONS!! to WØVXO/Herb for the 2nd Transatlantic QSO EVER from Wø when he QSO'd G3PU559 re 3 VFB!!! WØNWX has only other Wø-European 160 QSO on record. He also QSO'd G3PU 11 yrs ago!!- Wish could print all of WØVXO's letter/Space limited-however he hrd DL1FF569/0730z w/Horiz Doublet & NOT AT ALL w/Vert. Could have kicked myself for not putting up doublet sooner he sez. NOTE WØVXO & WØNWX both also hrd Novll by PY/Swl Rolf Rasp 0545z - FB!! p/1BU-2FBA,IU,FYT,GGL,TMG-K2BWR-3GQF-4VYB,KTA-Wa4FJ 8IRE,HRV,OOR,QWI,GHY-K8UNP-9PNE,CUY,HLY-K9GZK-ØVXO,IFH-VE1ZZ,2IL,UQ-3DU,AGX-1BB EAST:::: G3PU's signal was terriffic and outstanding, staying in from 0500-0800, peaking 579-averaging 559 dipping 229 with slow QSB and wkg quite few Ws includin WØVXO & W9PNE!!!!(CONGRATULATIONS to Ern also on his first QSO with VP8GQ this AM DL1FF was next-QSOing quite a few but not up to usual stbength. G6BQ,30QT,IGW,8JR also made QSOs. (The previous evening Dec15 G2PL569 wkd W3GQF and band was quite open half hr with fair signals) ---- Gsw1/QUEST hrd 1BB,ME w/30' piece wire at g level!! and nite before 1TX,ME-2FYT,KQT VFB!! ----Gsw1/Allen hrd 1BB,ME,TX,VP8GQ using series tuned "LOOPSTICK" 2/short 12" screened lead!!! Sone goings-eh??? say it helps out QRM/N -sumpin to try!!! ---- Gsw1/Iball hrd 1BB only Cdx bad Winds above storm force SOUTH:::: VP8GQ QSO'd G3PU/0507z CONGRATULATIONS Peter - W1 listened to both sides of QSO!!! Night before - 15th G6BQ also QSO'd VP8GQ for po FIRST - (On 15th QSO'd W9PNE and Brice's son W9CDP at 0400z) VP8GQ that is. ----- VP3AD/"Dad"hrd 2FYT569-K2BWR559-W9HLY459-W2FYT589-W1BB559-W1TX559-W8HRV459-W2GGL4 VE3AGX449 between 0300-0510z -----KP4AXU on 3.5Mc works 1.8 Mc Stns Cross-band -- PYswl/Rolf Rasp sends impressive list hrd 0500-0600z, VP8GQ589-W1BB589-2GGL589- 8IRE449-2FYT459-1ME459-2IU579-VE3AGX449-1TX459-K9GZK349-8WI239 Condx outstanding QRM/high-QSB/strong. Also 14th Dec hrd WØNWX569/0700,W6KIP589/0713 and 15th VP8G 338/579 QSO G6BQ-0515z WEST:::: No stations further West than W9/WØhrd and none reporting to W1BB NORTH:::: VO1BD heard clg DL1FF/0435z-believe no QSO. Dx-G3ORH,OQT,ERN,NVO,GGS,ROZ/a,RKA,RIQ,OFA,PWK,LIQ/a,OIN,KDV,MCY,PSB,PQA-6BQ,TK 8JR-ZL2BE-OH3NY-OK1MGmAEZmKSH,#CEA-GM3AEYpAWF-SVØWZ-GW3CBY NOTE !!!! During this week the band openedcup Dec 2nd and G3PQA put in a terriffic signal QSOing W3GQ W1BB,ME-2IU-UWD-8HRV. G6BQ QSO'd W1BB,TX - and W1BB QSO'd VP8GQ ((9PNE also QSO'd PQA)

Dec23:::: WWV sending "W-3", which is extremely poor Cdx - and so they were- not a trace of DX and all participants QRT early. Strange when morning before G3PQA put on such good show. Understand terriffic storm in G/land

NUCLEAR TESTING ON 160!!! Vs 160 Dx !!!

An extremely interesting letter from George Jacobs/W3ASK comments on our FLASH bulletin Nov.11 ---"I hve mentioned the possibility of A & H bomb explosions causi absorption in the ionosphere, with some friends of mine in this field. They are of the opinion that the eneggy contained in the many bombs exploded recently by both the U.S. and the USSR is still negligible to the amount of energy that would be necessary to induce a long period absorption effect. With Bomb-testing apparently halted, how have conditions on 160 been?" - (George, since testing stopped conditi seem to be coming back to normal for this time of year, but extremely slowly and v very spotty - whether there is any correlation or not is hard to tell - but some still feel that there is, others that there might be!!) Conditions hard to explain

FINLAND ON 160!!! ----- Here is another Country available - word from several sour OH3NY, GW8PG, etc - NO HAMS allowed 1820-1845k & 1915-1955k - 10 Watts. Our old frie Matti Paiavo/OH3NY, also OH2NB are on - Matti says "Cdx much poorer than 11 yrs ag sigs must go over Magnetic N Pole - will be active even though new QTH hr much poo for ant - 73 to all!!"

PROPAGATION FORECASTING ----- Fm WØTUT/Joe a most interesting comment --"Your coincidence of QSOs p3 Bull#1 62/63 reminds me paragraph June 62 RSGB as follows - ""IEE procedures 3/62 explains new method ionospheric forecasting being used by th D.S.I.R. based on identification of "Epochs" during which corresponding Cdx have prevailed past yrs, rather than plotting completely new forecasting maps as previo It has been found usually possible identify some period within the previous 10 yrs when almost identical radio condx occurred. The result of this new system, is a great increase in accuracy of predictions mdde several months in advance" - Tnx Jo Sorry abt ur S9 line noise washing you out on 160 Dx - Joe wants info on D.F. to find noise - can anyone help?? If so send Joe the references - Joe Check CST/ARRL.

ULTIMATE NAVIGATIONAL AID!!!!!"----- Fm Boston Globe 11/8/62 -"A new Navigational aid developed by John A Pierce, Harvard Scientist, (who also developed LORAN)-Named OMEGA - 8 times better than LORAN - Range 8000 Miles (Loran only effective 1000 Mi Three OMEGAs would replace 100 Lorans and give greater coverage -$10,000,000. cost per installation however. Expected operational not less 3 yrs nor more than 5 yrs- been in development 12 yrs. Funded by three Services through office of Navigations Research - apparently operates on higher Fy than LORAN (Dont know fy yet) -SOOOOoo maybe here's relief in sight from our LORAN QRM on 160 - We HOPE!! Anyone more dop on this????

HEARD ON 160 (Without mentioning any names) BY W1BB ----- "A fellow can stretch his ears -when digging for Dx on 160 & or any other band - but especia 160 -to the point where he THINKS he hears sumpin!! Did YOU hear ever so weak a si on 1827?? I thought I did!!! - Gosh, I doubt if my ears will ever go back into sha " - Now isnt it the truth???!!!! W1BB at least knows JUST what you mean, O.M.!!!

MARKER DHJ54/2 ----- Per G3PQA "Has packed in and is replaced by DHJ59/2 which is much weaker station - so if you dont hear DHJ - dont despair of Dx"- Tnx John !!

CYPRUS - Fm G2BVN/Steve comes word that 5B4PB is Famagusta, Cyprus & Now active - usually fm 0300-0450z 1801k alternate odd numbered days. Static high/but will TRY!

160 in FLORIDA ----- de W4XYT/Andy Florida Skip Ed. - "You said there is nothing keeping us fm at least listening in Fla to 160!! Oh YES there is Stew - LORAN!!!!! I'd like invite you Fla just ONE Night - ANY night and "Listen" to 160!! You proba couldnt stand the 40 to 60 DB over S9 racket fm our Hi Pwr LORAN Stns - Man it's JU AWFUL!!!! - lets hope however some day they'll turn off those QRM makers an allow us back on"!!! Tnx/Andy-real sympathies and hopes for a better future!!!

"FLORIDA SKIP" Mag/W4JO "Work For Joy Only", Frank, Advt Mgr, sends extremely FB comments --"160 been dead band Fla many yrs-missed band mucho- W4EB/W4JO organized "Miami 160 phone gang" in 1930-Frank used 201a Mod W/247-still has it -(In Attic)- in 1933 Mag "HARMONICS" born devoted 160/Phone/Hamming-Net met 8AM Sundays - so yo see 160 is vy dear to hearts of O.T.s arnd Miami - we used Long Wires/Low Pwr - Dx was'anywhere out of the state' tho I once QSO'd Europe with tt 201A - wl be list Transatlantics - If poss thru Loran -was THERE at the very FIRST & 2nd TRANSATLANT w/cans only " Frank O.O.T./Tnx Vy interesting Comments - Hope 160 back to Fla Soon

AMERICAN SHORT WAVE LISTENERS CLUB -----Jim Howard, Ed, says "Thinking of offering awards to ASWLC members who report overseas 160 meter Dx stations - might make it Club Contest" - this would help Dx stns a lot!! Tnx much Jim - Hope you do it!!!

SOUTH AFRICAN RADIO LEAGUE/Michael C Bosch/ZS2FM,East London ----- Writes " Our br[illegible]h has embarked on a project to revive interest in "Top-Band" in view of low M exp[illegible]ted during Min Sunspot Cycle. Following stns are on - ZS2CE,CD,FF,FM,GN,JH,KA KW,LC,CF,OL - some VFB/Most Xtal 1950k - ZS2FM has rock for 1987 - pwr 10w,198[illegible]kfr OK!!!! Julius Reitmayer/Club Station OK1KCI ----- says, trying hard for W contacts Requests contacts & Contestants especially for their "TP""TEST" 2nd & 4Th Mondays 160 - fm 1900-2100z - Exchange rpts RST plus QSO Nr ie 579013 - Call CQTP. One QSO pts Multiplier is No. QSOs during 1st half hour 1900-1930z. Send TEST Logs within 3 days to Central Radio Club PO Box 69 Praha 1 Czechslovakia - OK??? (Time NG W/Ve

MISCELLANEOUS TOP-BAND NEWS ITEMS -----G/swl/Norman Smith was season's earliest,la latest, most persistent SWLer!-His rpts kept up all summer! Oct6 band sounded STIN no activityNorm kept at it finally digging out W8HGW & VP8GQ fm mud 30th VFB!!! -- GW8PG/Gus,Oct14 h/G6BQ,3IGW,CQT,OIT,PQS,NQF,GSS,active/NoDx-Gus vy happy/Psed w/ne Friends/Activities 160-says "More Antenna dope & SWL rpts fm W/Ve,Pse"!!! DL1FF/Ar Per Gswl/Quest)who can be counted on to give 160 real working over, has VEE beam,5 high towards WEST being hrd terriffically loud in Europe-broke thru to W/VE 449 No DL9KP,1WW,6OS also active. -----EI9J/PADDY (11/6) Hopes be on soon - says "Regardi your saying that Trans-Equatorial Cdx best summer - I could agree - Over last 12 y Jun&Jul hv bn two best months in yr - QRN rarely high hr-we get no extremes - I am particularly thinking of 80/same goes for 160 Tnx Paddy/bonu ----- OY7ML/MARTIN, g spcl Lic operate 160 on 1825k during TRANSATLANTIC DX TESTS. First night on wkd 20 stns inc. many Gs, EI9J,DL1FF,GW3JI - Hrd a N African,could not raise-using 10w - Congratulations Martin!!!! ----- USA/SWL/Adrian Fallert/Hamilton Ohio taking part Dx logging per GW8PG/Gus who was hrd by him Sep30,0600z,449. This is VFB to have U SWL!!! -----Gswl/QUEST rpts DJ2KS,3KR,1FF,KB,7IL,1WW,6OS active 160-also NOV7th OY7ML "Exploded" on 160 wkg string Gs,DL1FF fr FIRST ever DL/OY 160 QSO!!! OH2NB w DL1FF Nov13 & G3MYI on CW&SSB for a 1st!!! 5B4PB emerges fm QRN now and then-TnxTo VP3AD/DAD wl be on regular TRANSATLANTIC TESTS only whenever poss. 1801k - can mak no special skeds at other times/time limited - had rig trouble/now fixed - Good ne G3OQT/Dick says "I want to cross pond SSB - Anyone interested please contact me fo series skeds weekday mornings-(So not QRM boys hr weekends/make myself unpopular!) OK SSB gang write R.F.McLachlan 27 Lake Rise,Romford Essex Eng - Dick nw has new ant bent 1/2 wave -has crossed pond 4 times since Sep30!!!'VFB!! -----PYswl/RolfRa doing FB listening 160 Hrd W1BB several times on "CQ DX TEST" calls & other ?????? weak signals - band bad PY land however NOV. ----- G2HKU/Ted has new QTH "Now actu above sea level" wl b on soon. -----G3CHN/Roger"Patchy season so far-G3FPQ wkd VP8 OH2YV,NB & PA[illegible]KN,SS active" Tnx Roger -----EL4A/Ken"It doesnt look like I'll ever on fm Liberia agn/confiscated all gear Sep27-wl be lvg spring-hope go some rare D Dx Country next to make up for lost time" Congrats to Ken&Traut on VFB yl Jr Op!! ExGD3MIK/Rex 160 Dxer nw VE2BKK is at Perry Iowa "Long live 160-wish I could be mo active" -----GM2BUD/Dave wl b on starting Jan 9 - His SON graduated "Assbciate"-Ro College Science & Technology & Wkg DeHaviland Aircraft - VFB Dave/Congrats-get him interested 160!!! -----ZS2FM&ZS6XQ&"RADIO ZS"Mag plugging 160 in S.A. - bound to h more ZS 160 activity soon!! -----G3PQA/John"Wl be glad make special skeds with any anytime" -----G2DCG/Ernest plugging 160 hard-wkd UA3NB,DJ2KS,OY7ML,GD3UB,OK1AA[illegible], DJ3KR,1KB,2KS -----PA[illegible]N anxious for Ws-Nuf Sed!! -----JIM/GM3IAA(ExVS1AA) wl hvo new mast soon & active 160 agn -----G3OIT/Keith wkd OH2NB,DL1FF& Others -----W1WY/ FRANK regrets time doesnt permit much time on 160-has 75A4-intends some activity- offers give good CQ Publicity to "HOT" 160 News - Tnx Frank -----W9PNE/Brice "Neve hrd cdx so poor-dec3-WWV/"W-3/4" for nearly 2 Mos mw-only few times up to N5-You h no idea hw difficult it is to wk 160 Dxfm out hr - Noise level low however - testi various ants/interesting & Confusing results -Wkd 80 exclusively during CQ/WW DX t after 6th Straight Certificate-Hope!! Congrats Brice!! ----K4KSY/Rev DraytonCooper Moved to Southport N.C.-on soon as can solve ant problem/little space-wl use new SP-600-JX-14 Rx - bonu Dray on 160!! -----W8ANO/Waldo ex avid 160 Dxer listens 160 on old BC/Rx but still no Tx-too bad!! ----W6KIP/Alex moans "Out here I dont know 160 could be worse!!" All I can hear is big juicy power leaks arnd clock" Sri/Alex

73 de Stew/W1BB and b.c.n.u. on 160 follows HAPPY NEW YEAR!

November 11, 1962

!!! FLASH !!!

MIDDLE NOVEMBER

!!!-- 160 METER DX report de W1BB --!!!

Dx in all directions since October 1st have been extremely poor, although some spotty DX has broken through to W/VE, namely, VP8GQ (Weak) - HR3HH good signal - and DL1FF weak and briefly for 20 minutes (No QSOs) using his Vee beam and calling CQ DX --- European "MARKERS" have been conspicuously absent except for the briefest moments.

This raises a question as to what has happened to 160 meter Dx. Up until October, it was our best year - with the earliest Transatlantic crossing on record and lots of Southern Summer DX including VP8, ZL, VK etc and apparently continuing to improve while the SUNSPOT Numbers increased, as expected.

However - rather suddenly from the first of October on, it has been almost as though a DX barrier had been thrown up, which even the strongest DX signals can not penetrate.

Many are wondering if the Pacific NUCLEAR tests could be responsible for upsetting the Ionosphere and casting a "blanket" over 160???????

All 160 Meter DXers are active however, and holding on hoping for the "Break" to come soon!!!!!

Most active W/VE PARTICIPANTS are: W1BB, TX, ME, UZ, WY-K1IMO-2EQS, FYT, IU, UW1 KHT, -3GQF, FBV, -4CXV-5KG, NU, SCT-6KIP-Wa6CDR-K7DEK/7, K7HDB-8GDQ, IQS, CIJ-K8IQR-9PT UKV-K9JDK, -ANWX, IFH, GBV, VXO-VE1ZZ, 3QU, DU, FF DX PARTICIPANTS are: G6BQ, G3ERI LIQ, PU, CHN, OIT, PQA, IGW-[redacted] PU- G5RI-G6GM-G8PG - GM3IAA - GM3IOS - GD3UB - HR3HH - VP8GQ - EI9J - UB5WF and others.

Stewart S. Perry/W1BB

(During Week of December 2 W1BB will visit KV4AA "Dick" in the Virgin Islands --To discuss possibility of 160 Meter DX operations from "YASME and future DXpeditions

1960/1961

DECEMBER 23 ~~4th~~ 16 and ~~17~~ ~~18~~
JANUARY 6 7 ~~8~~ 20 and ~~21~~ ~~22~~
FEBRUARY 3 ~~4~~ ~~5~~ 17 and ~~18~~ ~~19~~

During these periods, Extra Special efforts will be made to contact European, Asian, African and other 160 Meter DX Amateurs throughout the world, for general DXing, and also yo make observations, as a contribution to the study of PROPOGATIO conditions on these frequencies.

An active group of English and DX Amateurs, in Cooperation with USA Amateurs are behind this effort in carrying on these Tests, which have been a yearly operat activity since 1932. All 160 meter stations throughout the world are invited and also urged to participate in as many of these TESTS as possible!!!

Most W/VEs will operate 1800-1825 Kc Segment EAST of the Mississippie River, while those to the WEST will use 1975-2000 Kc, Clustering near 2000 Kcs Usually. DX will be found mostly between 1795 and 1850 Kcs, the best segment being 1800-182 Some DX will be found near 2000Kcs. DX stations should cover near 2000Kcs carefull for unusual and choice USA Contacts!!!

W/VE stations will call CQ DX TEST the first 5 minutes of each hour, and each alternate 5 minute period thereafter, Listening the second 5 minute period, and each alternate 5 minute period of the hour, for DX. The DX stations of course will do just the opposite. Thus, each station alternately calls for 5 Minutes, and then listens for 5 minutes throughout the TEST - unless working or hearing DX. Care should be taken particularly by all participants to carefully set their clocks to WWV, etc. to be sure of keeping in step with these 5 minute periods, and not cause QRM by transmitting during the listening periods. These calling and listening peri of course apply only when no DX is being heard or worked. When DX is coming throug the idea naturally, is to QSO regardless of the periods - BUT, here again, adheran to the 5 minute periods, even for a QSO, would help to minimize QRM.

W/VE stations should send their reports for the TESTS,of activity and unusual accomplishments to W1BB, Stewart S. Perry, 36 Pleasant Street, Winthrop, Mass., US for recording in the booklet " THE 160 METER DX HISTORY", and for transmittal through BULLETINS to A.R.R.L.-CQ-SHORT WAVE-WESTERN RADIO and other magazines, etc for publication. G,GM,GI,GW, Etc. should send their reports to G6QB, L.H. Thomas, DX Editor, Short Wave Magazine, Turkey Road, Bexhill or Sea, Sussex, England. --- OTHER DX, may send their reports to either of the above as desired. Do REPORT!!!

Remember!!! --- These are "Tests", and not CONtests -- for pleasure in an unusual operating activity, without "Competition", for a change, and to develope some propogation information as a worthwhile contribution to the art of Radio.

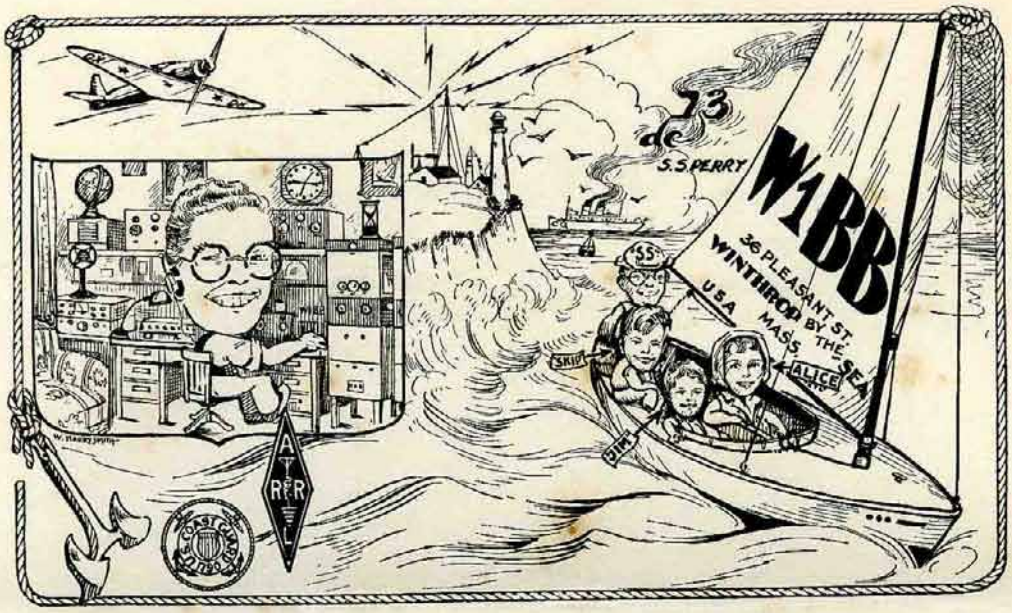
A.R.R.L. and CQ - DX CONTESTS

W1BB --- And many other W/VE, 160 Meter HAMS will carefully cover these contes on the same Operating schedule as above for the Transatlantics. W1BB invites any special schedule arrangements from any of the DX gang desiring to make a real try for an extra band multiplier!! AIR/MAIL your intentions to W1BB in this respect, in advance if possible, otherwise W1BB will cover the same frequencies as mentione in paragraph 3 on page 2 of this Special No. 1 Bulletin. CQ are proposing to run another 160 Meter CW Contest early in 1961, and it is hoped that DX stations will participate in this operating activity also. See CQ for further details.

W1BB 160 METER DX BULLETIN

W1BB --- Will issue this BULLETIN during the 160 Meter DX Season, as material is obtained, to the above/mentioned HAM magazines etc, via air-mail, so that they can publish from it, the important information on "TOP-BAND" activities. There is no mailing list to individials, although W1BB will oblige and send copies to those supplying S.A.S.E.s

Best of luck ALL - and again, 73 de W1BB/Stew.



W1BB ANNUAL 160 METER DX LETTER

Dear O.M.:-

At the start of another Season, W1BB sends GREETINGS with 73!!, and the hope t you will take an important part in the 160 meter DX activities!!!

Working DX on 160 is challengine, and extremely interesting. Conditions such a statics, BC harmonics, QRM, Loran, QSB, noise, all require extreme patience, with perseverance, a top-notch station, and A-1 operating ability. There are SPECIAL, TRANSATLANTIC and WORLDWIDE 160 Meter DX "Tests", and the ARRL and CQ, DX Contest to participate in. (See below in this bulletin). There are many "FIRSTS" yet to b Many unusual conditions to observe and cope with. One gets anew, the thrill of the early Amateur Pioneer days when HAMS were first relegated to "200 meters d down", and Paul Godley made the FIRST Atlantic crossing via Amateur Radio. In addition, you will many friends among the finest and most sincerely devoted HAMS you would ever care to meet -- these "Top-Band" boys !!!

Your participation here will also help to promote Amateur occupancy of the ban and assist in keeping this 160 meter operating priviledge. In addition, you will help to develope PROPOGATION information of real value to the art.

Through these BULLETINS issued during the season by W1BB, reports on your activities and accomplishments, and all information of 160 meter interest, will b sent to QST-CQ-SHORT WAVE-DX-WESTERN RADIO-VQM-DXing HORIZONS-FLORIDA SKIP-Radio Propogation Editor/CQ, magazines, etc., for publication. Be sure to keep W1BB wel informed of all your 160 Meter exploits and results, etc.

In case you personally are not particularly interested in 160 Meter operation, "Test" schedules, DX Contacts, or unable to work this band - PLEASE pass this BULLETIN along to someone who is. Also, W1BB would appreciate if you would send t calls of all HAMS in your Country who might be interested in 160 Meter DX, and ab to work this band. Please post this BULLETIN in your Club, and give 160 meters as much publicity as possible!! Particularly as to DX!!! THANK-YOU, very much indeed!

W1BB 160 METER DX ACTIVITY

W1BB --- Will be on regularly EVERY Sunday Morning, (Baring unexpected sickness business demands or Emergencies), from 0500-0730 GMT, calling and listening for alternate 5 Minute periods, November through March, looking for 160 Meter DX Contacts!!! Please be sure to give W1BB a call!!

W1BB --- Will transmit, Mostly on 1802½ or 1807½ Kc. However, 1817½ and 1822½ will also be used at times depending on conditions, QRM, Interferance, etc.

W1BB --- Will NOT work V/VEs during DX schedule hours - unless something of an urgent nature comes up, or band conditions are useless for DX - will Concentrate/L on

W1BB --- Will generally follow the regular 160 meter "Test" procedure. When no DX is being heard or worked, W1BB will call, CQ DX TEST, the first 5 minute period of every hour, and each alternate 5 minute period throughout the hour. W1BB will listen carefully for DX stations, the second 5 minute period of each hour, and each alternate 5 minute period throughout the hour.

W1BB --- Suggests that DX stations call W1BB on 1797½ - 1802½ or 1827½ Kcs, as these are the most QRM/free spots, usually. These SPOT frequencies will be covered with EXTREME care by W1BB at ALL times. The band 1800-1830 will be covered careful too, and frequencies up to 2000Kcs checked regularly.

W1BB --- Welcomes, and invites special schedules with any DX station that desires to make contact - at any time of year, and on any special, (Authorized), frequency designated by the DX station. Send AIR/MAIL letter to W1BB, with full details on the proposed schedules, and also any schedules on HF bands that may be necessary to make full arrangements or exchange of reports etc.

W1BB --- Appreciates REPORTS on his signals, from any station, at any time of year, with full particulars - whether you make QSO schedules or not. This includes SWLs, some of whom have submitted MOST accurate and valuable data. These reports assist in propogation studies being made. Please AIR/MAIL them weekly. Postage gladly reimbursed if desired. Special reporting FORMS supplied on request. All reports acknowledged. Please DO report!!! The more the better!!!

W1BB --- Will operate mostly from the HOME QTH, "Winthrop by the Sea", BOSTON, USA, with 200 watts, a 260' doublet antenna and 75A4 receiver. Some operation will be from W1BB/1, either at the WINTHROP YACHT CLUB, with quiet location and special antenna entirely over salt water!!, or from "Villa Mon Repos", at Harrington, Maine 300 Miles Northeast from Boston, a quiet country location, with 260° long "Vee" antenna - 260' on a leg.

W1BB --- Will mail a handsome appropriately lettered "AWARD" Certificate, to all DX stations working W1BB on this band, if they request it. In addition, for DX stations where their W1BB/QSO is a "FIRST" recorded "Top-Band" QSO between that Country, and USA, an extremely SPECIAL and attractive lettering job, in color will be done on the Certificate by a Commercial Artist, to Adequately and Especially commemorate the event. This Certificate is in effect a "KING-SIZE" QSL, measuring 10" x 14", beautifully colored, on heavy durable white paper, suitable for wall mounting or framing. DX SWLs who send in 7 or more Verified reports on W1BB's signals, or have in the past, may also have a Certificate upon request. DX stations who have worked W1BB in past years, and any previous "FIRSTS", may have one of these Certificates, back dated to cover the event, if they request it. Sixteen of these special "FIRST", Colored certificates have been awarded thus far, and in each case the recipient has written appreciatively advising that it was framed and/or posted conspicuously in the shack!!

W1BB --- Hopes that you will join in on the fun, and the challenging work of QSOing DX on 160 - and enjoy yourself and help others to make that much desired contact!! Write W1BB if you need further particulars. Keep W1BB fully informed of all 160 meter activities possible as suggested above, for the BULLETINS and reports on your doings to HAM magazines for publication, as their Editors decide.

W1BB --- Says, 73!!! HAPPY HAMMING and b.c.n.u. on 160!!!!! --- de Stew/W1BB

ANNUAL TRANSATLANTIC and WORLDWIDE "TOP-BAND" 160 Meter DX TEST !!!!

These special "TESTS" will be held this Season on the following SUNDAY Morning from 0500 to 0730 GMT, in accordance with the listings given below:---

Balun and Transformer Core Selection

[Home] [Up]

Related articles at

Balun Test contains model of "perfect" dipole currents.

Sleeve Balun shows how a sleeve adds impedance, useful for VHF and higher baluns

Receiving Common Mode Noise shows how lack of a balun can contribute to system noise (it applies to transmitting antennas as well)

Longwires, Verticals, and Baluns shows how unbalanced antennas can have similar problems

Transmitting baluns on testing transmitting baluns

Occasionally errors are made regarding core selection. This especially includes baluns, where on occasion some very strange ideas surface. One rather odd but somewhat popular idea is that adding a mixture of core types will allow both high power operation and high choking impedance in baluns by slowly reducing current through a balun. Other misleading claims are that extreme values of core ui, such as values in the 10,000 or higher range, are necessary on 1.8 MHz and higher. Other ideas tend support use of excessively low permeability cores for the same application.

When I recommend a core, the material selection is always based on actual measurements with proper test equipment on a bench as well as in the actual end-application.

Core Material

I mainly use 73 material for receiving applications in ***LOW POWER*** applications between .1 and 30 MHz. 73 and similar core materials generally minimize the turns count required without inducing excessive loss. One of the best indicators of correct core selection is looking at the turns required. You'll notice most of the transformers I use have only one or two turns for every 100 ohms of impedance. My 75 to 450 ohm Beverage transformers, for example, only require two-turn primary and 5-turn secondary windings. ***A low "turns count" is a good indicator the correct core size and***

core material is being used.

For high power applications at HF it is often necessary to use lower permeability cores. There are two reasons for this:

- Lower permeability cores generally are available with higher curie temperatures. They operate at high temperatures without losing their magnetic properties.
- Lower permeability cores have higher Q (lower loss tangent) at a given frequency. This means a larger part of the impedance is associated with *lossless* reactance rather *dissipative* resistance. They turn a smaller percentage of power into heat, and that is very important at high power levels.

Permeability changes with frequency. As frequency is increased from zero eventually core impedance peaks. Above the frequency where impedance peaks the impedance of the core (and the ***effective*** permeability) actually decreases.

A downward slope in permeability with increasing frequency is useful for controlling impedance in broadband transformers, but we should be careful to avoid excessive slope. Excessive initial permeability can easily move the operating area too far out on the downward slope of impedance.

A transformer or inductor operating on the downward slope of a high ui core requires extra turns to maintain critical impedance and often requires more turns than a lower ui core. The upper frequency limit will decrease, and this may reduce useful bandwidth in the desired frequency range.

Using excessive initial permeability means winding becomes more tedious (it takes more turns). The wire has to be smaller and more fragile to fit a given core window. Temperature stability is often reduced while losses increase over an optimum core material selection. In addition, stray capacitance increases needlessly, reducing bandwidth and increasing unwanted stray coupling.

Do **NOT** pick cores solely by considering initial mu, since that value is taken at dc. You should consider ***characteristics measured at the operating frequency!***

Always remember this general guideline. Les wire length (as long as winding impedance is sufficient) results in better transformer bandwidth. The best designs place maximum conductor length INSIDE the magnetic core window, and minimum conductor length OUTSIDE the core window.

Heating

At higher power levels, it is necessary to move to lower loss tangent and higher curie

temperature materials like 65, 61, or (in extreme cases) 43 materials. Even a fraction of a dB loss produces significant heating in small cores when power level is in the kilowatt range. *The loss DIFFERENCE in non-resonant applications between lower and high ui ferrite cores isn't significantly different, but* ***heating*** *can be much less!*

We often assume heat means a core is very lossy or is "saturating", but **t**his often isn't true. We must consider the power level, duty cycle, and ability of the core to dissipate heat and look at the full picture.

Very small cores, such as small thin .5 inch diameter cores used on bead-type choke baluns, can only dissipate a fraction of a watt in open air. It sometimes helps to put temperature in perspective by visualizing how hot a 60-watt light bulb runs in normal operation. When we consider the core's size, it usually has significantly less surface area than the bulb. The core also has poor thermal conductivity, and is often stuffed in a container preventing any type of air circulation.

Consider the construction of a typical bead balun, enclosed in PVC and heat shrink tubing. As little as 20 watts dissipated out of 1500 watts can produce damaging heat in tiny beads enclosed in a PVC tube. 20 watts out of 1500 is less than 0.1dB loss, yet it overheats the core!

The problem is almost always a heating problem, and not a core-loss problem. It is almost never core saturation, unless the core is subjected to very low average power and very high peak power levels. It is best that we worry about heat and the number of turns we use, not actual power loss, when selecting a core.

Core Style

Soft-iron cores (soft magnetically) increase inductance because they increase *flux density* near a conductor for a given current. With only a small amount of flux "concentration", there can not be a large increase in inductance or impedance. We need a significant increase in flux to have a significant increase in impedance.

The area outside the core window does NOT have a closed magnetic path surrounding the conductor. ***The presence of the core has a minimal effect on impedance of any conductor area outside the core window.*** Most of the flux from external wires is in air, rather the core. With only a portion of the flux surrounding the outer conductors cutting the outer layers of the core, the ***useful*** impedance contribution of wire outside the core window to system impedance is minimal. Conductor length outside the core window mostly adds unwanted stray reactance and leakage flux. If we MINIMIZE the wire length exposed outside the core, and we generally have a more effective inductor, choke, or transformer.

This effect can be easily conformed in a simple experiment using an antenna analyzer. Connect a short wire across the output of an analyzer, and measure the impedance. Lay a core against the wire,

and observe the very small impedance increase. Now pass the same wire through the core center, and observe the large impedance change. This illustrates why the winding's wire length on the outside of the core is wasted, mostly contributing to undesired effects.

Core Dimensions

The area inside the winding-window of a soft-iron (soft magnetically, not physically) core is cut by all of the flux lines, and this area has a very large effect on impedance. The core concentrates the magnetic flux surrounding a current-carrying conductor into a very small area, and the thickness of the core moving away from the area of the conductor very rapidly has less effect.

- The additional impedance caused by placing a core over a conductor or conductors is almost entirely proportional to the core's internal length (window depth) paralleling and surrounding the conductor or conductors.
- The core diameter or radial thickness only has a small effect on impedance.

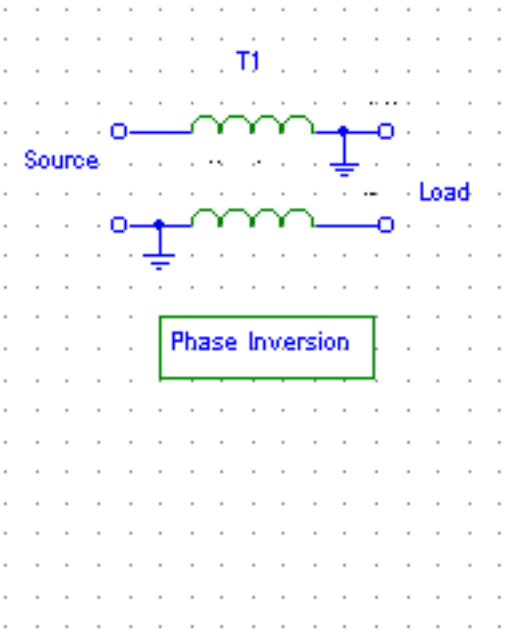
Doubling the core area **parallel** with a conductor roughly doubles winding impedance. The same is NOT true for an increase in core wall thickness, core thickness barely affects impedance.

I prefer binocular cores for most low-power applications and side-by-side stacks of cores (making a large "binocular core") for high power broad-band applications. This type of core arrangement almost always minimizes the amount of conductor hanging "outside the window". With very little conductor hanging "outside" the core window, there is less "needless" wire adding undesired stray capacitance and series resistance. For a given core material and impedance, conductor length can often be reduced to about one-third of a similar impedance choke (or transformer) using a conventional single-hole core or single stack!

Low Power Measurements

Phase Inversion and Choke Baluns

Some of my receiving system designs use phase-inversion transformers. Phase-inversion transformers are identical to (and interchangeable with) choke baluns or line-isolation transformers. For HF receiving applications, 73-material binocular cores are wound with six passes of #26 twisted-pair enameled wire. I use Fair Rite Products 2873000202 cores (about 1/2 inch square and 1/3 inch thick 73 material).

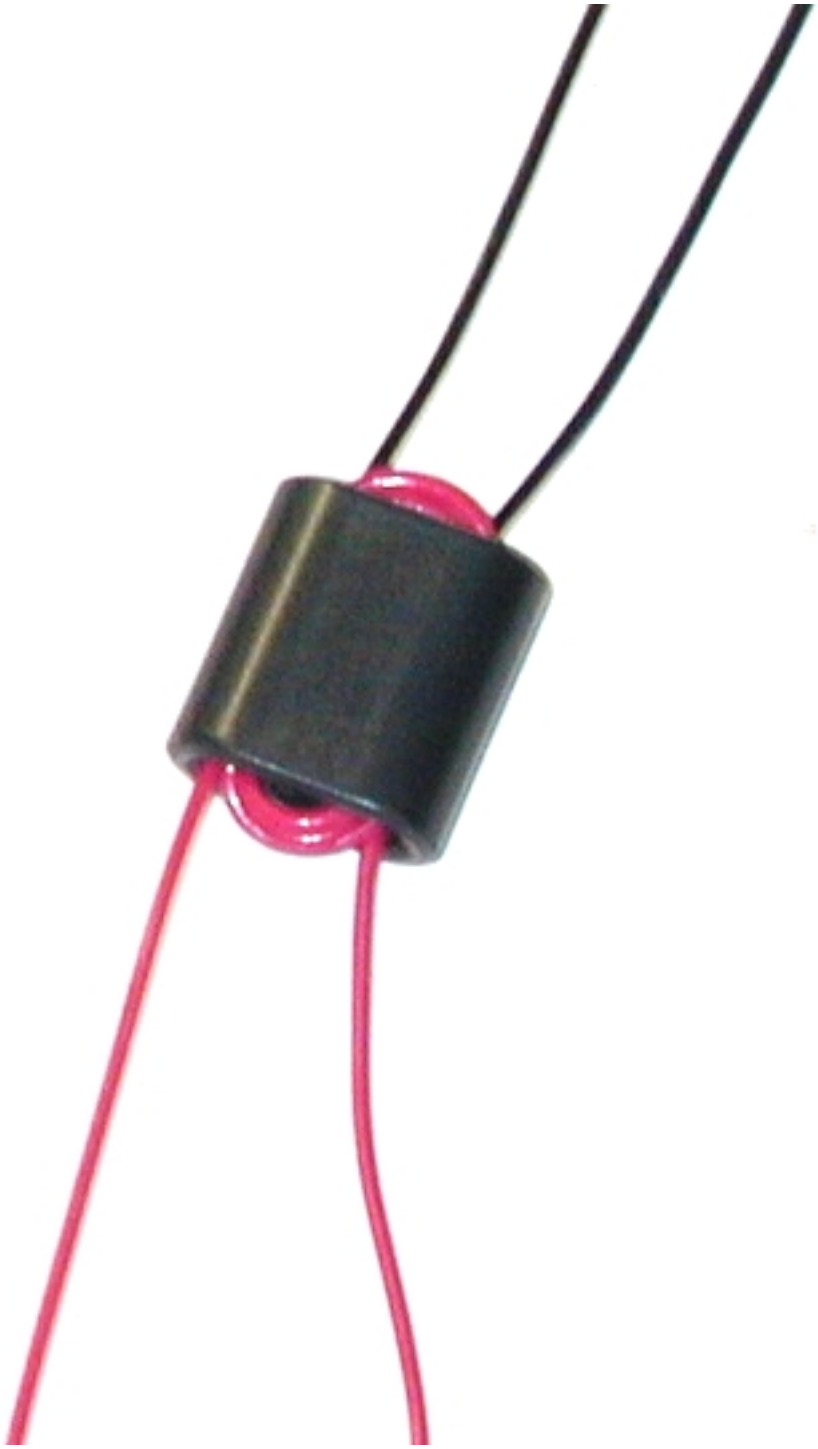


Here are measurements of a sample inversion transformer at 2.5MHz using accurate (and fairly new) commercial equipment:

Load Value (ohms)	Loss (-dB)	Phase Error (degrees deviation from 180)
100	~ 0	<1
33	~ 0	-1.2
10	0.2	-2

This shows inverting transformer construction is good, since even a 10-ohm impedance load works well!

Beverage Matching Transformer



Sometimes I use designs for a long time, and forget how I decided they were OK. I recently received an e-mail questioning the number of turns in my Beverage transformer design, so it seemed like a good time to re-confirm the design.

I retested a 2:5 turn ratio transformer using a single FairRite Products 2873000202 core (about 1/2

inch square and 1/3 inch thick 73 material) two different ways on a generator/ network analyzer/vector impedance test set.

Total loss of two back-to-back transformers was .84dB at 1 MHz increasing, not decreasing, linearly to .98dB at 30MHz. The actual transformer loss would be .42dB at 1MHz increasing to .49dB at 30MHz.

Doubling turns increased the attenuation slope. While 1MHz loss decreased to .69dB per pair, 30MHz loss increased to 1.21dB. This was for a PAIR of transformers connected in series to make a 1:1 transformer. This of course removes mismatch losses, so it is twice the real transformer loss. Actual loss would be .35dB @1MHz increasing to .61dB at 30MHz.

Measuring a second way, I terminated the transformer in 470 ohms. Loss measured .65dB at 1MHz when mismatch loss was included. Since receivers have wide ranges of input impedance, any mismatch error might help OR hurt actual system loss. Factoring out mismatch loss the second measurement indicated about .53dB 1MHz loss with the original 2-to-5 turn transformer and .43dB with twice the turns.

Every measurement has tolerances, and the two different methods do provide different losses because measurement errors affect results differently. Still, it is safe to say doubling turns has a negligible effect on 1MHz loss (which is around .45dB).

Balun Test

[Home]

Related pages:

Sleeve baluns Balun and Transformer Core Selection

Impedance and SWR Test

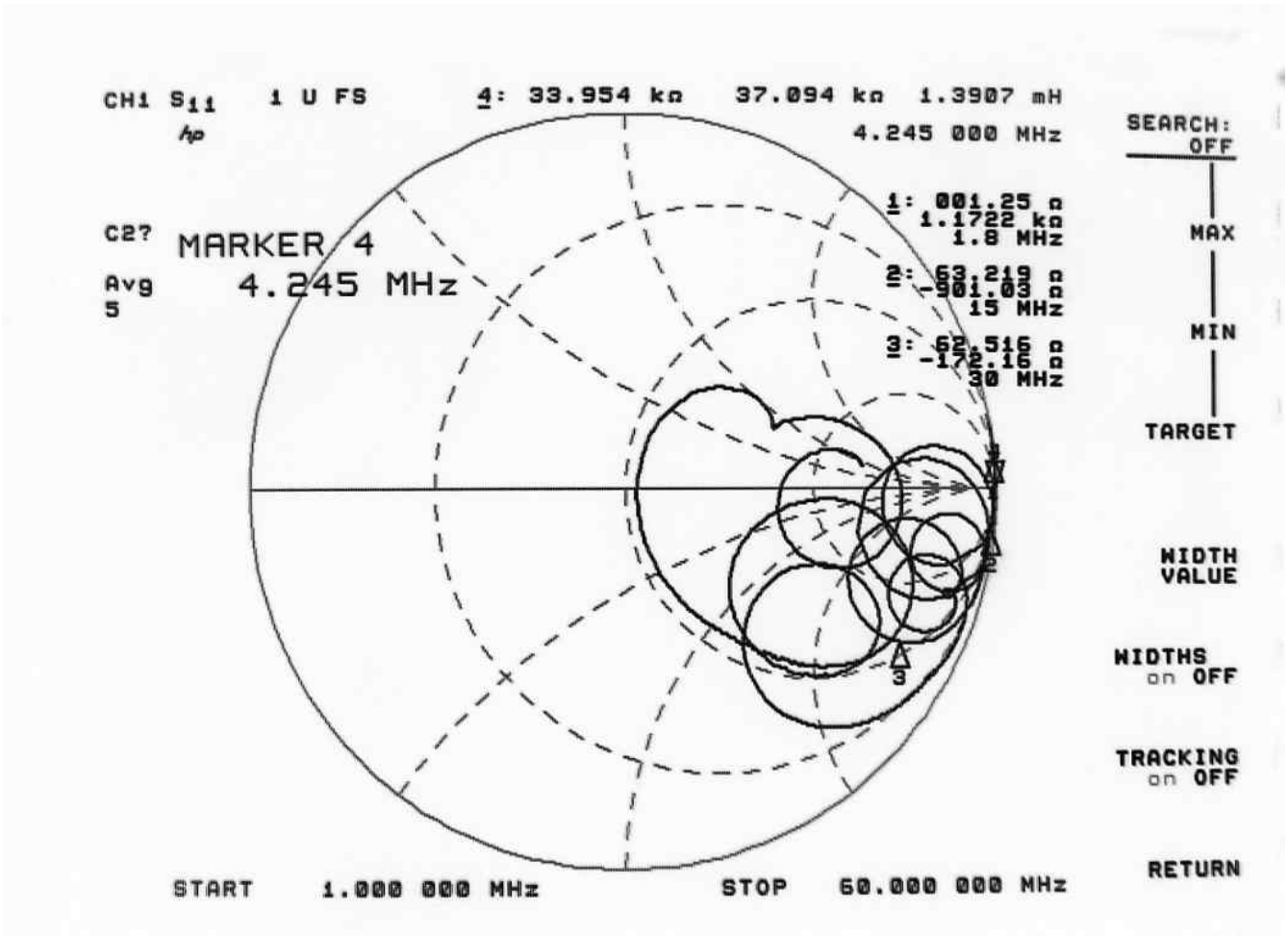
This page shows measurements of various baluns and how various baluns compare.

Choke Impedance

This data shows the common mode impedance of the balun. In general, the highest impedance at the operating frequency or over the operating frequency range is desirable. This impedance isolates the antenna from undesired signals on the feedline shield, and prevents antenna terminal voltage from exciting the feedline with unwanted currents. Common mode impedance is directly related to the care in design and construction.

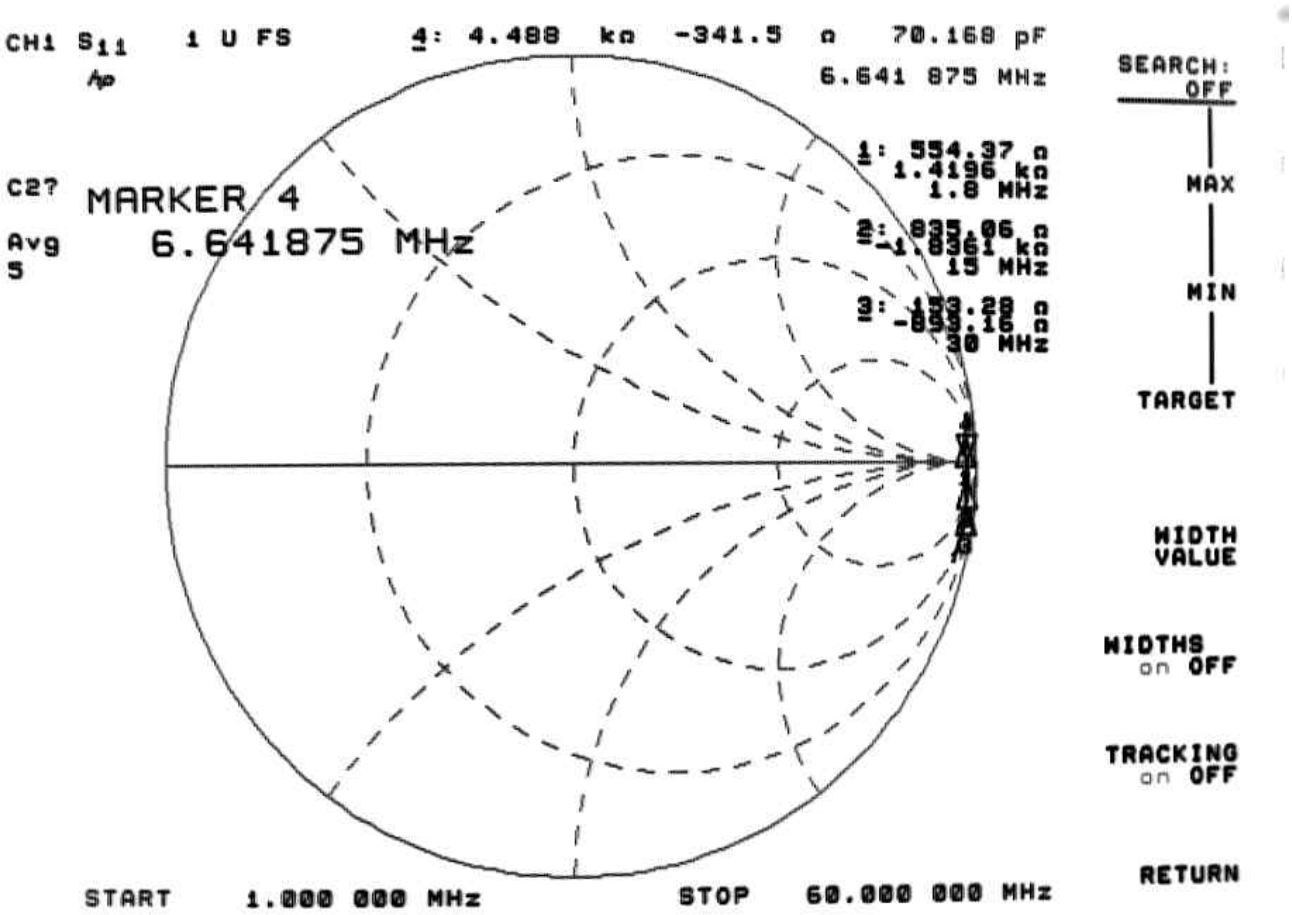
Pay particular attention to the impedance peak in air-core baluns. For narrow-band applications they make excellent baluns.

Unfortunately common-mode impedance is all over the place, as this Smith Chart plot shows:



The air-core balun is good only for a three or four-to-one frequency range, unless you pick a winding style and size that places unwanted series resonances outside desired bands.

In contrast a good core-type balun looks like this:



SWR

The lowest SWR is desirable, although any mismatch can often be compensated by adjustment of antenna dimensions. This SWR mainly comes from incorrect wire impedance inside the balun. It may be caused by excessive length of internal leads, or incorrect cable or winding impedance inside the balun. It generally is a construction related problem.

Measured SWR and Choking Impedance

The following data is measured using a currently certified network analyzer with low capacitance test fixture:

Centaur	DXE	W2DU(1)	W2DU(2)	Force 12	Scramble	Solenoid	W2AU volt

Choke Impedance								Note 2
R+X@1.8	84 129j	554 1.4k j	378 617j	230 325j	169 286j	1.67 245j	1 +1.18k j	.488 5j
R+X@15	3.76 2.7k j	835 - 1.84k j	727 -611j	761 -10j	883 - 105j	11.97 - 850j	62 -895j	1.36 42j
R+X@30	143 - 729j	153 - 893j	284 -440j	610 -296j	538 - 381j	73 162j	68 -168j	8.2 68j
Max Z@F	17 MHz	6.65 MHz	7.16 MHz	15.3 MHz	13.24 MHz	6.42MHz	4.25 MHz	60 MHz
R+X at max Z	5.87k - 943j	4.5K - 340j	1.3K -13j	770 -20j	914 2.25j	42.7k 0j	34K 37K j	75 286j
Min Z@F						27.68MHz	11.7Mhz	
R+X at min Z						10 -2j	198 -252j	
SWR								
F SWR=1.25	6.8 MHz	65 MHz	21.15 MHz		20.2 MHz	note 1	note 1	20.9 MHz
1.8MHz	1.07	1.02	1.03		1.03			1.43
15MHz	1.58	1.04	1.17		1.18			1.2
30MHz	2.16	1.08	1.37		1.39			1.35

note 1: SWR not measured because construction and cable type affects SWR

note 2: This is a W2AU voltage balun. It is only shown as a example of poor shield isolation offered by voltage baluns if the antenna is not perfectly matched to the balun with the feedline exiting the balun at right angles. This type of balun is unsuitable for non-symmetrical systems such as off-center-fed antennas, verticals, or antennas with the feedline paralleling the antenna (even at a fairly large distance).

The W2DU baluns were manufactured by Unidilla. (1) is a Maxi balun and 2 is a 10-40 meter model.

The DX Engineering balun is the dipole balun type DXE-BAL050-H05-P

The scramble-wound choke was about 20 feet of RG8X in a six-inch diameter "bundle".

The solenoid balun was about 60 feet of RG-8X on a 4" PVC thin wall drain pipe coated with rubberized roofing tar.

Power Dissipation and Feedline Common-mode Current Estimates

Note: This section revised 1/2/2003 to correct model error. Please report any other errors to me!

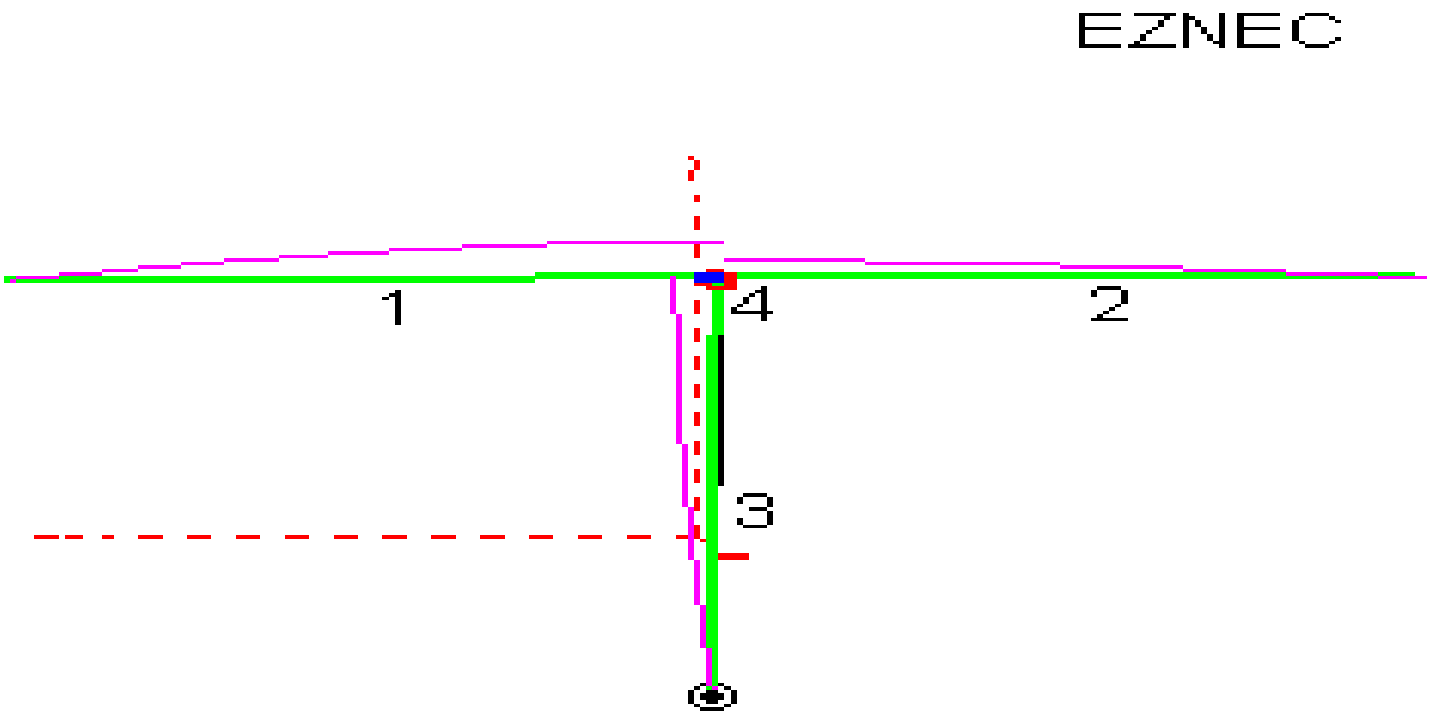
Balun power dissipation is estimated using Eznec to simulate a perfectly balanced dipole.

Here is a copy of the model used:

160m balun EZ

Please be aware I made no special effort to create a "bad antenna" other than I intuitively understand what the worse case condition of feedline length would normally be and I selected that length. I dropped the wire representing the feedline vertically from the center of a perfectly balanced dipole, and made that wire 1/4 wl long.

Here is a view of the model with no balun:



SWR is 1.46:1 power is 1500 watts

Currents at 1500 watts are approximately:

5.65 amperes into wire 1

2.63 amperes into wire 2

3.73 amperes into wire 3 (coaxial cable shield)

Using this model (a 135 foot high 160-meter dipole) we can add each of the balun impedances in the coaxial cable shield and estimate feedline current and power dissipated in the balun:

Centaur	DXE	W2DU(1)	W2DU(2)	Force 12	Scramble	Solenoid	W2AU
0.90 amp	0.12 amp	.25 amp	.47 amps	.57 amps	.87 amps	.16 amp	.4 amps

69 watts	8.5 watts	25 watts	51 watts	55 watts	1.3 watts (26w TL loss)	.03 watt (74w TL loss)	34 watts

From this we can see the following:

- Adding more beads is very inefficient. **W2DU(1)** has about twice the beads as **W2DU(2)**, yet it has 53% of the current and 49% of the power dissipation! This does increase power rating by a factor of four, but it is still too low to prevent balun heating.
- The **solenoid** has (by far) the lowest choking or balancing power loss, but it is 60-feet of RG-8X wound on a 4" diameter form in a single layer. It adds transmission line loss of 74 watts, but since the area is so large it will not overheat.
- The **scramble wound** balun has inadequate impedance since wire length is only 20 feet. It has low loss, but it really isn't acting like a balun (at 6.5 MHz it would be super, having 42K ohms of impedance).
- Of the baluns above, only the DXE, scramble wound, and solenoid would not be overheated in normal operation for continuous Morse CW transmissions.

Perspective of Heat

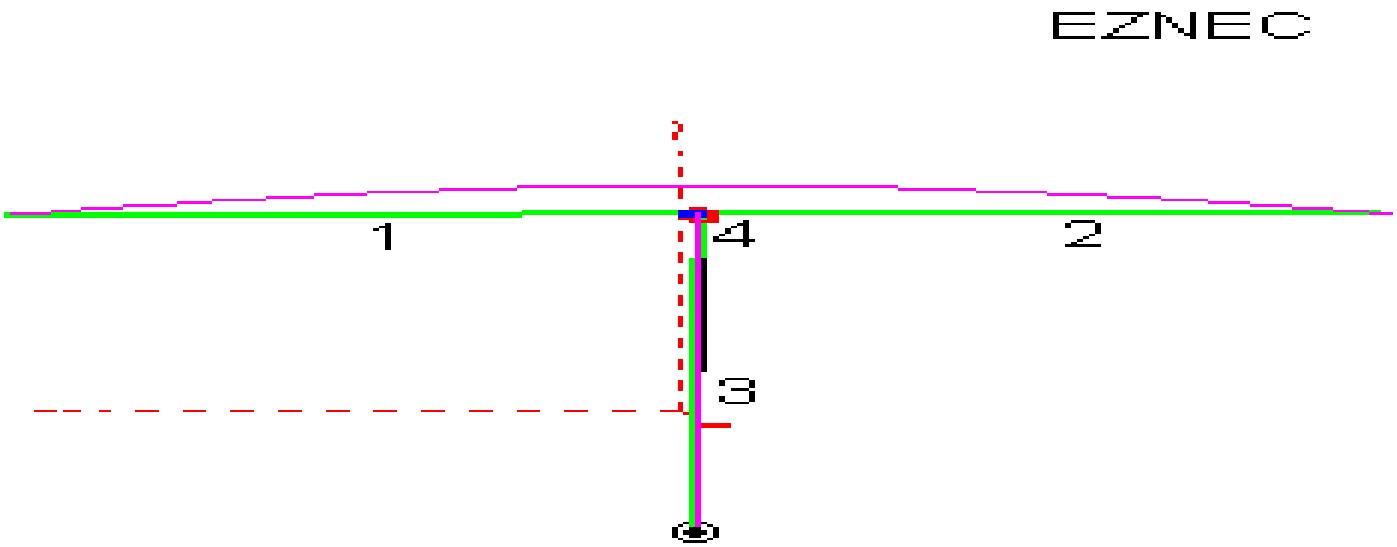
Think about the heating this way. Imagine you had a 60-watt light bulb. Nearly all of the applied power is turned to heat, and the surface area of the bulb and conduction through the base radiates that heat. Would you hold a 60 watt light bulb?

Now picture a balun core with a surface area a fraction of the size of the light bulb. This core area is enclosed in a case that often has poor thermal conductivity.

- The large air-core baluns mainly produce heat from transmission line losses.
- W2DU and other style baluns mainly have CORE losses. Transmission line losses are negligible since the transmission lines are very short.

Currents

Every ampere of current not going onto the cable shield goes to the dipole's shield-fed leg! The total is not the exact sum, because of phase differences. Here is a view of currents in the dipole with the DXE balun:



Currents are:

Wire 1 4.52 amperes

Wire 2 4.52 amperes

Wire 3 .12 ampere

Remember power radiated increases by the square of current. The feedline shield current is now .12A compared to 3.73A without a balun! The feedline radiates about .1% of the power it radiated without a balun. While pattern distortion on transmit may not hurt, the feedline radiation probably aggravates RFI and allows noise to couple into the antenna when receiving.

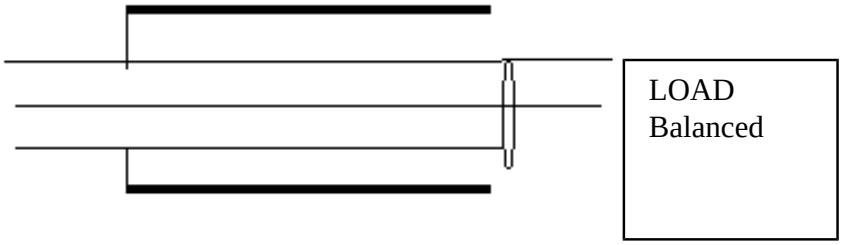
Remember the model is worse case in the NO BALUN condition. This does not mean every system or most systems will be this bad. This example was only intended to show how bad balance can be and how much power baluns (even with a matched load) can dissipate!

I'll add more data as time permits.

Sleeve Baluns

Sleeve baluns are normally used at VHF and higher. As a general rule, they are not practical at HF.

The sleeve is connected like this:



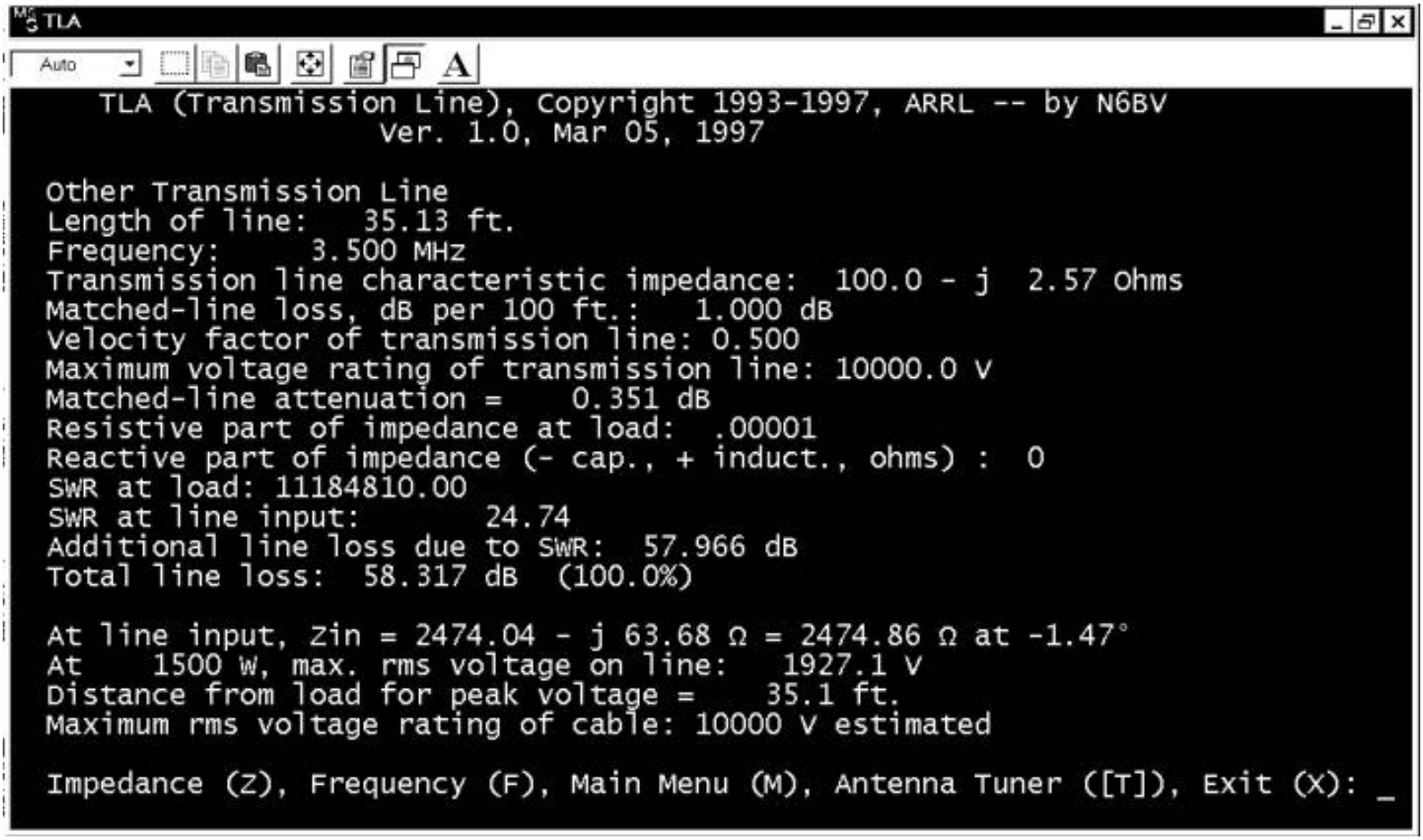
The left hand side of the drawing, once past the sleeve connection point, is UNBALANCED. The right hand side is balanced. The ***choking impedance*** of the balun is the impedance between the outer sleeve and the outside of the shield of the coaxial cable inside.

The choking impedance represents the same type of impedance we would have from any common-mode choke, such as a coil of cable or a string of ferrite beads would produce. A sleeve balun only works where it is ¼ wl electrical length, or odd multiple thereof.

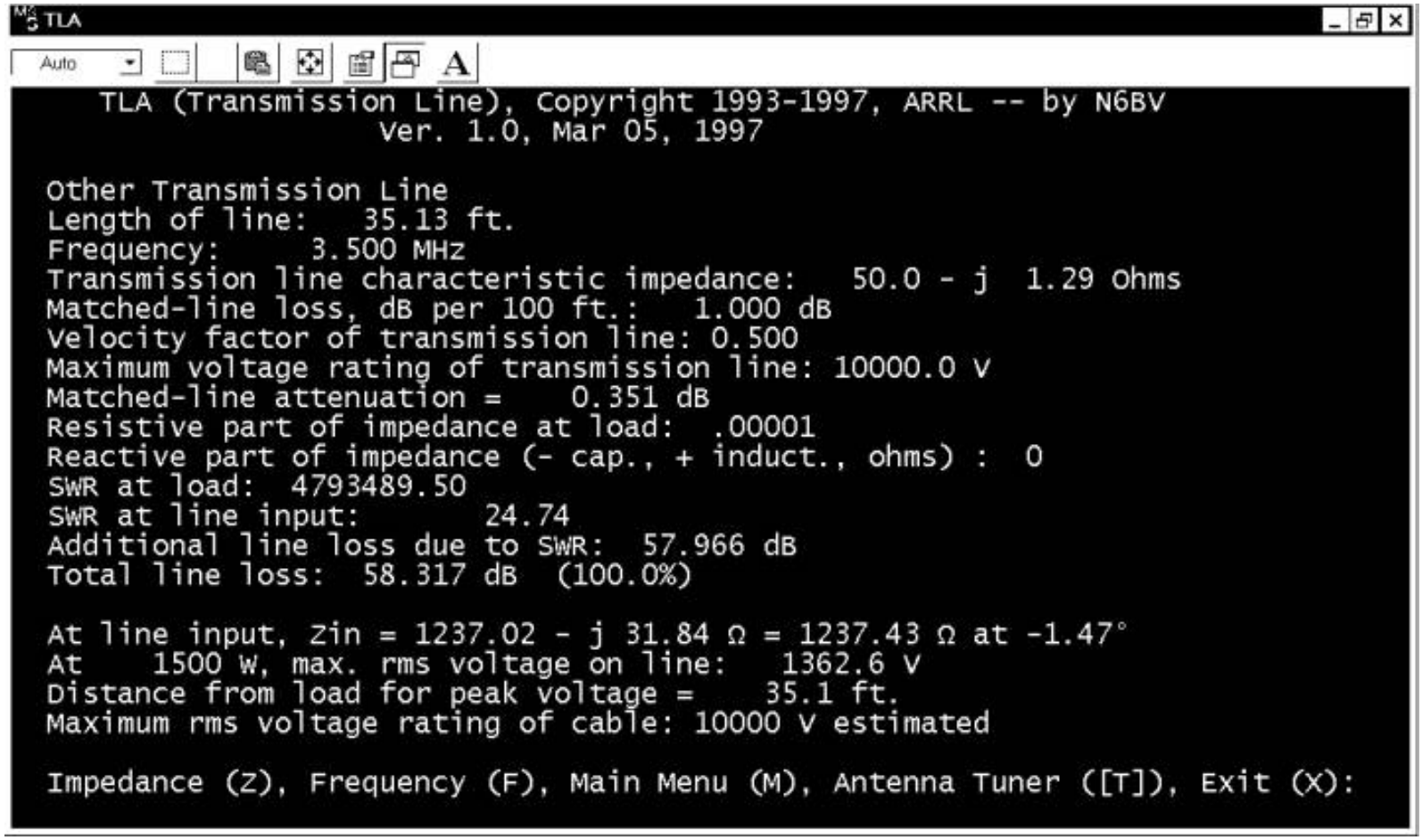
Obviously we want the highest possible choking impedance, because the balanced load terminal voltage on the shield side to unbalanced (shorted end side) voltage and impedance sets the current though this balun (as it does with any choke balun).

We can use TLA (free with the purchase of ARRL Antenna Books) to calculate balun's choking impedance.

Below is a calculation of sleeve balun choking impedance with 100-ohm Zo 1dB loss sleeve. I used a .5vF inside the sleeve.

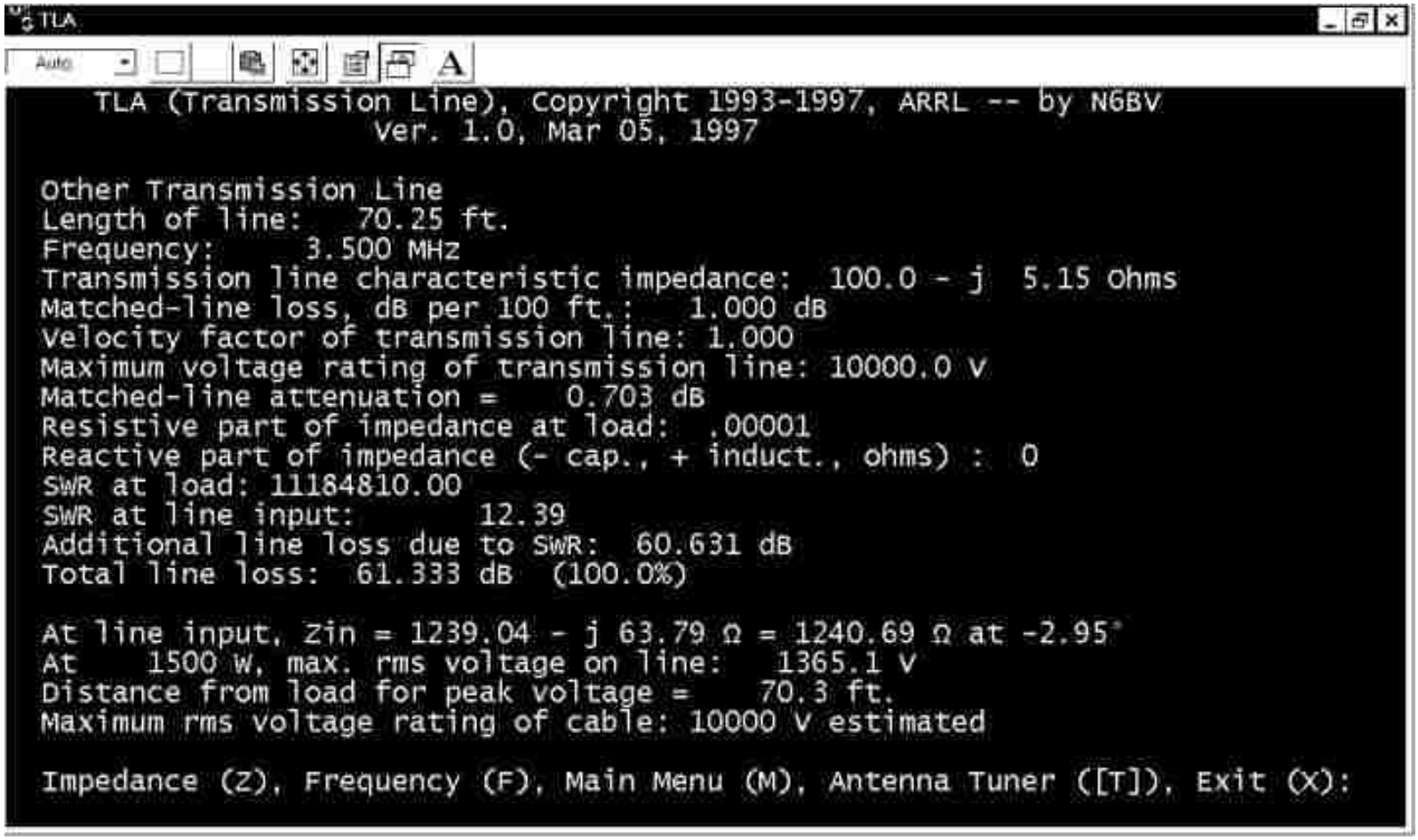


Balun choking impedance 50-ohm Zo sleeve and .351dB internal sleeve loss (sleeve 35 feet long and matched loss 1dB per hundred feet).



Impedance is now 1237 ohms compared to 2474 ohms for 100-ohm sleeve Zo.

As you double sleeve Zo, all other things equal, choking impedance doubles. What happens if we only change loss in the 100-ohm line? We can do this either by doubling the loss per hundred feet or doubling Vf at the same loss, making the sleeve twice as long (and twice the loss).



Now we have 1240 ohms with a 100-ohm Zo sleeve. The reason we have less Zo is the attenuation or loss is twice as high. If you look at Matched Line Attenuation above, it is now .703dB versus .351dB in the 0.5 Vf case.

From the above, we observe the following characteristics in a sleeve balun:

1.) The highest possible choke sleeve impedance (largest ratio of balun sleeve diameter to outside of transmission line) is desired. We won't have a good balun if the choking Zo (ratio of sleeve inner diameter to coaxial shield outer diameter) is small.

2.) The balun requires the lowest possible loss over the length of the sleeve. It forms a transmission line from the inside of the sleeve to the outside of the coax. The coax jacket is a dielectric, so we need to keep a lot of air inside of the choking sleeve or the coax jacket will increase loss and reduce impedance, both being very undesirable.

3.) The velocity factor of the sleeve, based on the dielectric between the sleeve and the shield of the coaxial cable we are

trying to balance or choke, is very important to length of the sleeve.

The following construction guidelines apply:

The cable should have a good low-loss jacket or a very large air or low loss dielectric gap between the shield and the sleeve. Since energy is normally confined to the inside of a coaxial cable manufacturers are not concerned about jacket losses. They use outer materials with long life, not low RF loss. It is advisable to use a filler material with a high volume of air to maximize sleeve impedance and minimize sleeve losses.

It is also advisable to use the largest practical diameter sleeve with the smallest diameter coaxial cable inside to maximize choking impedance.

The sleeve length has to account for velocity factor of the sleeve, since the sleeve forms a coaxial transmission line with the outer conductor of the coaxial cable it is intended to choke or decouple.

Common-Mode Noise

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Related articles at

Balun Test contains model of "perfect" dipole currents.

Sleeve Balun shows how a sleeve adds impedance, useful for VHF and higher baluns

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Common Mode Noise Isolation

Common-mode currents can be detrimental to antenna system noise or directional performance. A quick look at systems using common-mode currents demonstrates how effective they are at causing radiation. If common mode currents can radiate effectively; they can also transfer unwanted signals and noise ***into*** our antennas when receiving.

Many antennas actually function because of common mode currents. Two popular examples are the CFA and EH antennas. Both become significantly poorer radiators if common-mode currents on feedlines are eliminated. Another recent example appearing in Antennex's compact antenna articles was a thick stub "vertical" with no counterpoise. You find an example on the baluns and verticals page of how poor some large grounds can be.

A receiving example of an antenna that works because of common-mode excitation is the "Snake"

antenna. This system accidentally (or intentionally) induces common-mode on a cable shield in order to receive signals. The entire shield picks up signal, the Snake is simply a reverse-fed random wire lying on the ground.

There are many examples where designers intentionally use common-mode currents. Examples are found in textbooks, such as the "Antenna Engineering Handbook" by Jasik on and around page 22-6.

The antennas below, copied from Jasik's textbook, outline the derivation of a skirt collinear antenna from a simple feedline with the open end terminated by a "stinger".

22–6 VHF AND UHF COMMUNICATION ANTENNAS

Two or even three skirts may be added to the mast as shown in Fig. 22-5*c*, but reduced excitation of the lower sections diminishes the effectiveness of each additio skirt.[6] Thus, while the multiple-skirt coaxial antenna resembles a collinear arra in-phase half-wave elements, its gain is not as great. Typical gain values are ta lated in Table 22-2, but since no test data are available for this type of antenna, values given are merely engineering estimates.

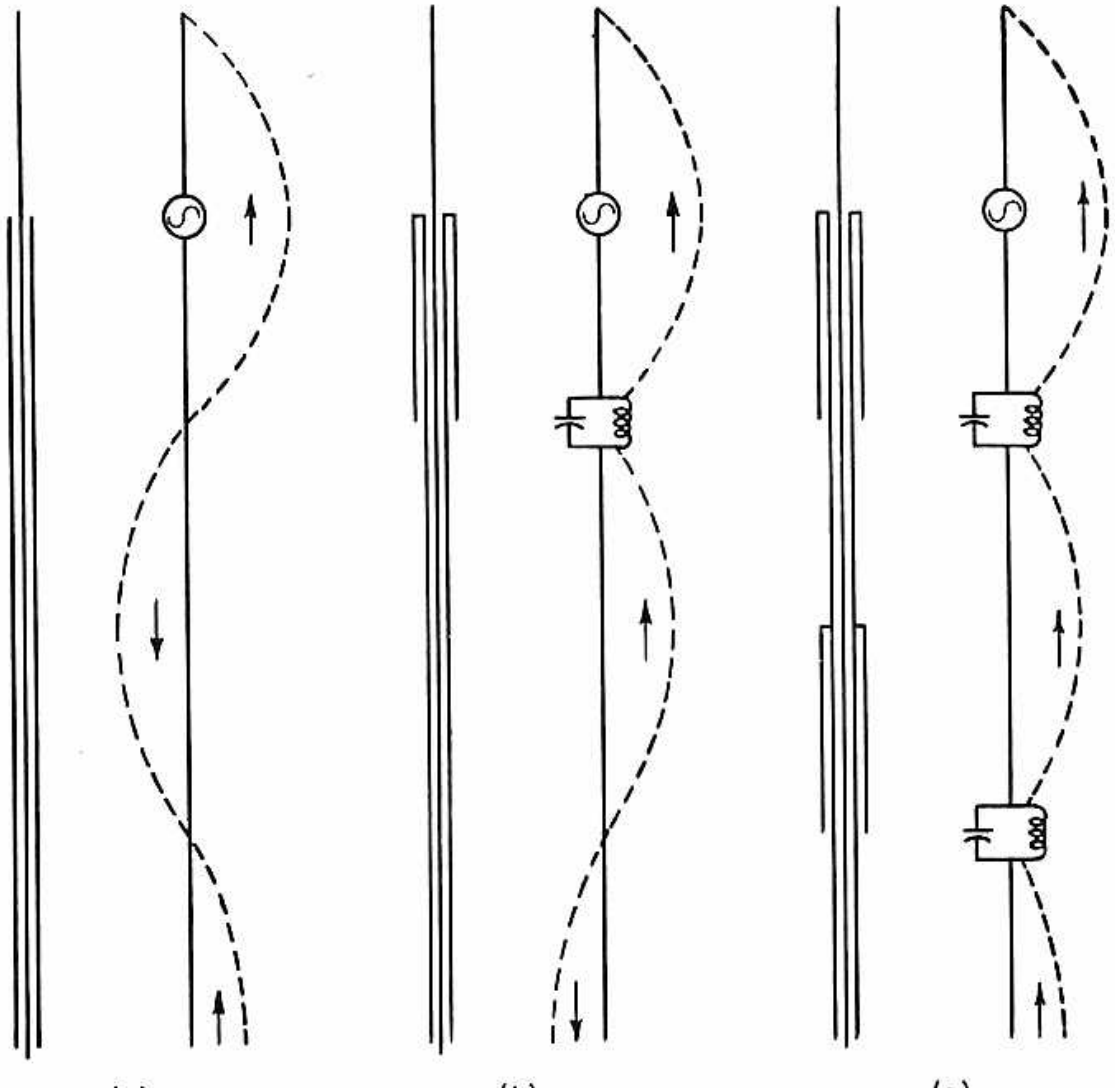


FIG. 22-5 Evolution of multiple-skirt coaxial antenna.

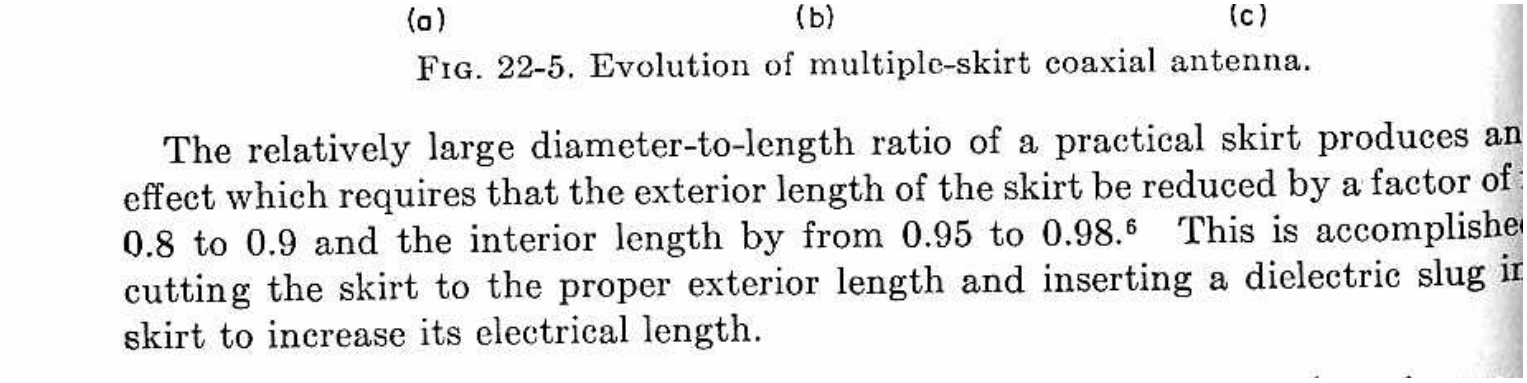


FIG. 22-5. Evolution of multiple-skirt coaxial antenna.

The relatively large diameter-to-length ratio of a practical skirt produces an effect which requires that the exterior length of the skirt be reduced by a factor of 0.8 to 0.9 and the interior length by from 0.95 to 0.98.[6] This is accomplished cutting the skirt to the proper exterior length and inserting a dielectric slug in skirt to increase its electrical length.

The center conductor termination in these drawings could easily be a ground rod (in the case of a Snake) or an antenna like a Beverage or loop. The termination does not have to be an "open circuit" 1/4 wl stinger that intentionally radiates!

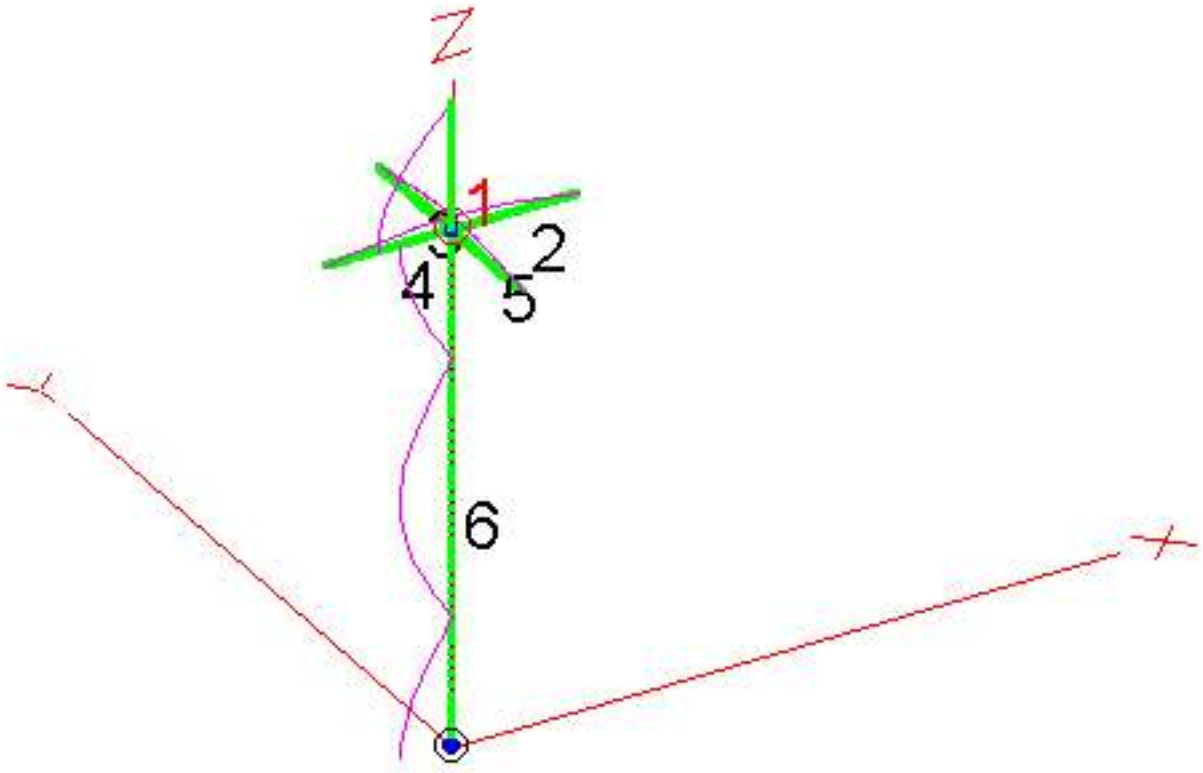
Looking at (a), we find by hanging any low impedance on the end of a coaxial cable the shield is excited by common-mode current.

The electrical equivalent is just as if the transmitter or receiver (generator symbol in the drawings) is located at the end of the shield. This causes the outside of the shield to act like a longwire antenna.

Unless the coaxial shield connects to a zero resistance ground, current with flow on the shield. Looking at (c), we find even multiple sleeves appearing as parallel tuned high-impedance circuits do not fully decouple a shield! It takes grounding and series impedance to do a good job.

Common Mode Currents and Receiving Antennas

Analyzing our antennas, we often forget grounds are not perfect. We make assumptions that four radials, or worse yet two radials, form a perfect groundplane. Even a groundplane antenna many wavelengths from earth with four radials has considerable common-mode currents on the feedline. Consider the following model of a "perfect" Ten Meter groundplane using four perfectly horizontal 1/4 wl radials spaced every 90-degrees with a 1/4 wl feedline hanging vertically and attached to the radials. The main element current was set at 100.



EZNEC ver. 3.0
Groundplane 12/14/02 6:20:40 PM
---------- CURRENT DATA ----------
Frequency = 29.95 MHz.

Wire No. 1 Main Element	100
Radial wires	19.71
Feedline shield at GP	69.96
1/4 wl from GP	2.12
1/2 wl from GP	71.3
3/4 wl from GP	.70557
At ground end of feedline	71

A glance at radial current shows the bulk of ampere-feet (ampere-feet, or current over spatial distance, determines E-M radiation levels) is on the feedline shield, not the antenna! Radiation from the feedline would be severe, yet most amateur antenna designers claim with only four radials, or worse yet two radials, no balun is needed! The claim that four radials makes a "perfect ground" is false.

Why do we depend on a simple ground rod with 50 or more ohms RF resistance to clamp a coaxial cable shield to ground?

Receiving Systems

Admittedly the above antenna is a worse-case example of feedline length and grounding, but even the best cases could cause problems. A best-case system might be "nearly perfect" when transmitting, but it could be a disaster receiving when even minor amounts of conducted noise are present on the station ground. Noise paths exist through station wiring. Only the shunting impedance of ground connections and series impedance of the feedline shield prevent excessive unwanted noise ingress at the antenna feedpoint.

Very small levels of conducted unwanted noise often go unnoticed in large high-level transmitting antennas. Noise ingress would not be an issue if local noise levels on power lines are very low, especially if the antenna has substantial common-mode feedline noise rejection.

If a feedline is very long and lies directly on or is buried in the earth, ground losses can attenuate conducted noise or unwanted common-mode signals. Unfortunately, we almost never know if the feedline shield is contributing noise, because it is nearly impossible to measure the common-mode noise contribution of the feedline!

Measuring Common-Mode Noise

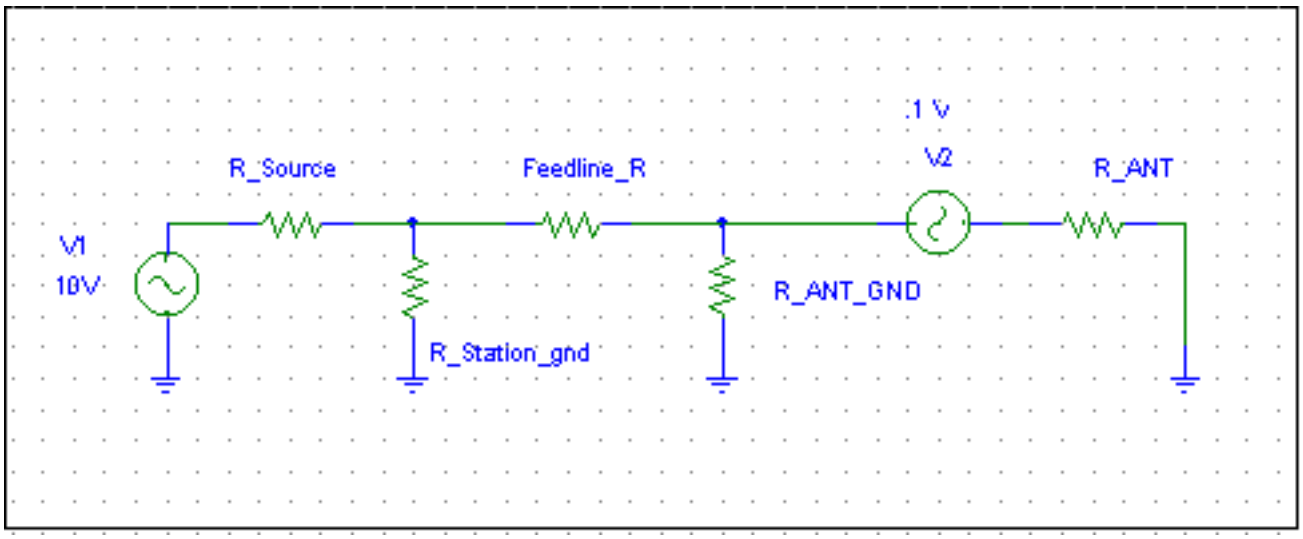
We sometime hear suggestion that we test a system for noise ingress by disconnecting and replacing the antenna with a dummy load. This idea actually has no theoretical foundation at all. The dummy load would change system common mode greatly.

The only real test would come from a dummy load with the same connections and impedances (both differential and common mode) as the actual antenna. In other words the test load has to be the actual antenna to keep feedline common mode ingress the same! Obviously this is a useless test!

The best approach is to use preventative measures in initial system design and installation. Quite often the cost of being safe is less than a few percent of the initial system expense.

Analyzing Systems

This circuit is simplification of typical common-mode paths in Beverage, EWE, and other similar antenna systems:



R_Source and V1 represent the source creating voltage across R_Station_Gnd, the station's ground impedance.

Feedline_R is the equivalent series-impedance of the feedline shield.

Current through the feedline shield path develops a voltage across R_Ant_gnd, which represents the earth connection ground impedance at the antenna.

V2 is a voltage source representing desired signals, while R_ant is an impedance representing the sum of the coaxial differential input impedance presented to the antenna (from the desired signal path into the coax) and the actual antenna impedance.

Using the circuit below, we can find the attenuation. Assume:

R_source is 90-ohms

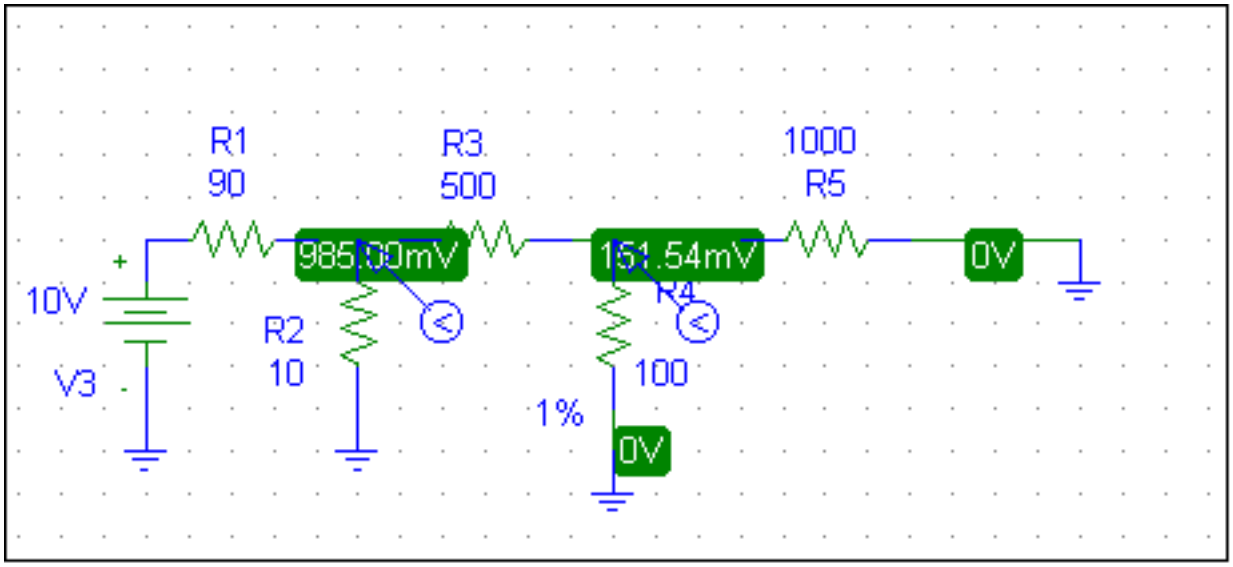
R_station_ gnd is 10 ohms

R3 (the coax shield) is 500 ohms

R5 is the combined series resistance of antenna impedance and impedance presented by the feedline matching system, is 1000 ohms

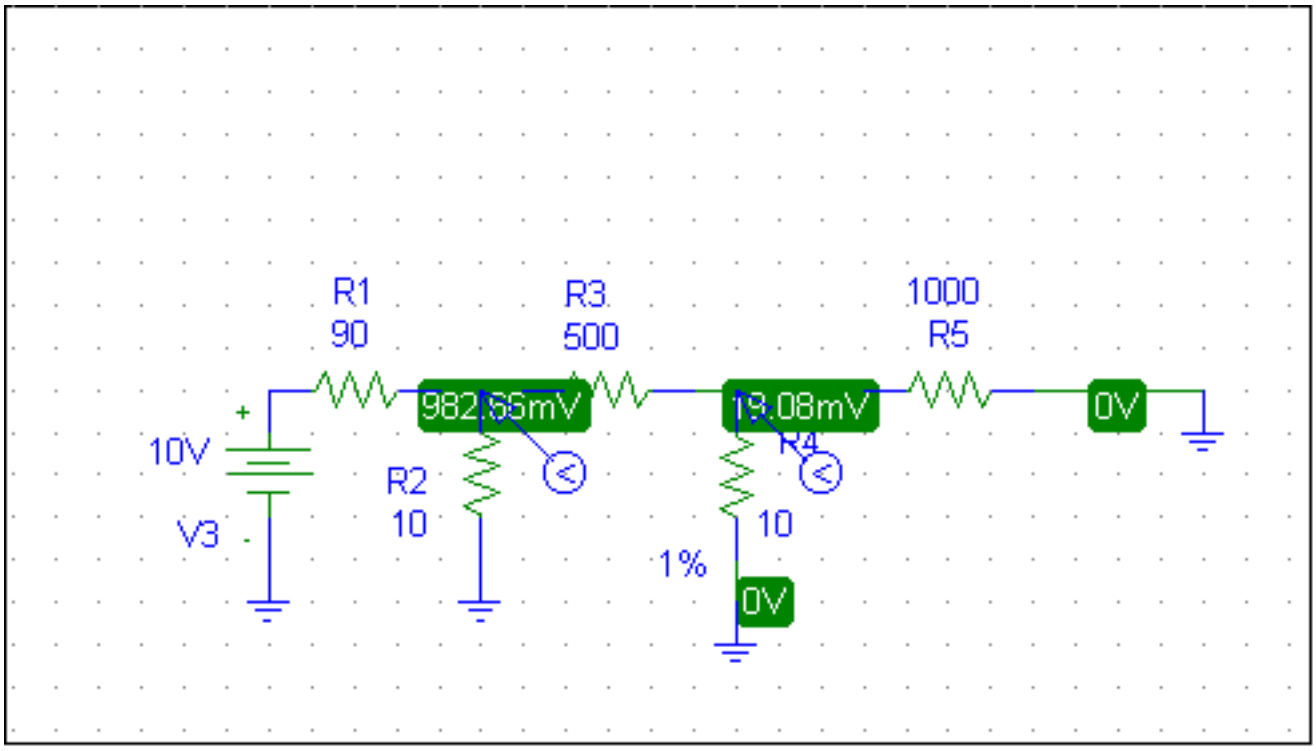
In a typical system where a single six-foot or deeper rod (the earth's skin depth prevents deeper ground rods from decreasing resistance substantially) is driven into typical soil, R_ANT_GND will typically be between 40 and 120 ohms, assume 100 ohms.

We have the following results:



Using the model above, only ~1 volt of common-mode voltage across the station ground results in .152 volts driving the feedline ***exactly as a signal from the antenna would***. Path attenuation from station ground to the feedline's differential input at the antenna is 20log 151.5/985 or **16.26dB.**

Changing the ground resistance to 10-ohms results in:



19.1/982.7 or ***~34dB attenuation*** of common-mode noise. Increasing R3 by adding beads has a similar effect. If R3 is effectively made ten-times larger, attenuation is in the 30dB range.

Obviously it takes a combination of reducing ground resistance and/or adding series impedance on the cable shield to significantly isolate any low-noise receiving antenna from conducted ground noise over

the feedline's shield.

We sometimes observe much less noise on transmitting verticals after installing a large effective ground system. Decreasing ground impedance at the antenna reduces common-mode excitation of the antenna feedpoint and reduces noise ingress, although adding a feedline choke would sometimes help.

Solutions

A typical isolation scheme would be to use an isolated primary and secondary in the matching transformer, and ground the feedline shield some distance away from the antenna's signal ground. This will introduce several thousand ohms of reactance in the common-mode signal path, as well as provide another path to earth for common-mode noise.

Another method, in cases where the feedline can not be isolated through a floating primary in a matching transformer, is the use of multiple independent ground rods with a series of choke baluns between each. This forms a multi-section pi attenuator, making even modest choke impedances effective. As an additional benefit, lightning paths are disrupted by this method.

Summary

Noise contribution can vary with time. A receiving antenna's ground connection resistance varies with soil moisture, and sources of noise come and go. As noise levels and grounding changes noise contribution as a ratio to antenna noise will change. The fact we can not readily measure noise contribution by substituting dummy loads further complicates the issue. Real systems are vastly more complex than the simple analysis above.

Since we can't easily measure noise contribution, we shouldn't take chances. It makes no sense to gamble that unwanted signals (from wrong directions) or noise are so low that they will never contribute to noise in a special antenna installed to reduce noise and interference.

While isolating feedline common mode effects from the antenna and antenna's ground may not reduce noise, isolation can generally be achieved at virtually zero time and material cost. With the low cost of prevention in mind, it is shortsighted at best and foolish at worse to not isolate a feedline shield from any low-noise antenna's signal ground path.

Follow these rules:

1. **With small magnetic loop antennas, make sure the antenna is properly balanced**
2. **Lay feedlines on or bury them in the ground so earth losses reduce shield current**
3. **Be sure shield connections are good**

4. **If noise levels are high or antenna sensitivity is very low, isolate the feedline from the antenna as it approaches the antenna by using choke baluns**
5. **Ground the feedline a few dozen feet away from the antenna**
6. **Avoid autotransformers. Instead use isolated primary and secondary winding transformers**
7. **Use an independent ground on the antenna. Do NOT connect it to the coaxial cable shield**

Noise

Also see: Receiving Antenna Design

To see world QRN (delayed): Lightning world map

To see my local QRN (nearly real time): Lightning USA Map

My local noise level quiet night from NE is -127dBm, 350Hz BW, from pair of ~800ft broadside Beverages ~375ft spaced.

The noise that limits our ability to hear a weak signal on the lower bands is almost always an accumulation of many signal sources. Below 18 MHz, the noise we hear on our receivers (even at the quietest sites) comes from terrestrial sources. Receiver noise is generally a mixture of local groundwave and ionosphere propagated noise sources, although some of us suffer with dominant noise sources located very close to our antennas.

Our locations fall into three basic "radio" categories that may or may not be related to our actual communities:

Note: noise levels quoted in this text are the average of three independent studies by Bell Labs, FCC Land Mobile Advisory committee, and the Institute for Telecommunication Sciences. Rural data are actual measurements of summer noontime and winter midnight noise at my location, several miles from high voltage transmission lines and far from any industrial or suburban populations.

Urban

In urban-type noise situations, noise arrives from multiple random sources through direct and groundwave propagation from local sources. One or more sources can actually be the induction-field zone of our antennas (in most cases the induction field dominates at distances less than 1/2λ). Urban locations are the least desirable locations because typical noise floors average 16dB higher

than suburban locations. There is often no evidence of winter night noise increase on 160 meters, since ionosphere-propagated noises are swamped out by the combined noise power of multiple local noise sources. Much of the noise sources are utility distribution lines, because of the large amount of hardware required to serve multiple users. Other noise sources are switching power supplies, arcing signs, and other unintentional man-made noise transmitters.

Suburban

Suburban locations average about 16 dB quieter than urban locations, and are about 20 dB noisier than rural locations. Noise generally is directional, arriving mostly from areas of densest population or the most noise-offensive power lines. Utility high-voltage transmission lines are often problematic at distances greater than a mile, and occasionally distribution lines can be problems. The recent influx of computers and switching power supplies has added a new dimension to suburban noise.

There is often a small increase in nighttime winter noise at exceptionally quiet suburban locations. This increase occurs when propagated terrestrial noise equals or exceeds local noise sources.

Rural

Rural locations, especially those miles from any population center, offer the quietest environment for low-band receiving. Daytime 160 meter noise levels are typically around 35-50 dB quieter than urban, more than 20 dB quieter than suburban locations. Nighttime brings a dramatic increase in low-band noise, as noise propagates in via the ionosphere from multiple distant sources.

Primary local noise sources are electric fences, switching power supplies, and utility lines. I can measure a 3 to 5dB daytime noise increase in the direction of two population centers, Barnesville (population 7500, distance 6 miles) and Forsyth (population 10,000, distance 6 miles) Georgia.

Typical daytime noise levels, measured on a 200 foot omni-directional vertical, are around -130 dBm with a 350 Hz bandwidth (noise power is directly proportional to receiver bandwidth) . Noise power increases about 5 to15 dB at night, when the band "opens". As in the case of suburban systems, directional antennas reduce noise power.

Nighttime is the "big equalizer", reducing the advantage of location as distant noises increase with improved propagation.

Polarization

Noise is generated by randomly polarized sources. Noise polarization is ***filtered*** by the method of propagation.

Noise arriving ***from the ionosphere*** is randomly polarized. It arrives at whatever polarization the ionosphere happens to favor at the moment. It has the same ratio of electric to magnetic fields as a "good" signal.

Noise arriving ***from sources within a few wavelengths*** of the antenna is also randomly polarized. It has NO dominant field, it can either be electric or magnetic field dominant.

Noise arriving ***from ground wave sources*** some distance from the antenna is vertically polarized. This is because the earth "filters out" any horizontal components. Horizontal electric field components are "short circuited" by the conductive earth as they propagate, and are eliminated.

To hear a demo of noise and directivity, go to the DX Sound page.

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MFJ-1025/1026

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Noise Phasing and MFJ-1025/1026 Data

The original concept of the 1025 is based on a bridge phasing system. Many phasing boxes, such as the QRN Squasher, do not rotate phase. They are based on the concept that a 180-degree inversion allows rotation of phase when mixed through a potentiometer, but that does not occur. The only thing mixing a 180-degree shifted phase signal with another base phase signal does is vary the amplitude, and produce an abrupt 180-degree flip at the zero-crossing point. Some people have copied the flawed QRN-Squasher concept, and offer similar products.

I recently reviewed another phasing system design (on paper) which is implied to be the "ultimate". It uses a simple R/C phase advance circuit, not a bridge. The maximum phase rotation of such a system is limited to less than 90 degrees. With 180-degree or zero-degree inversion in a later stage, that allows only a 150-160 degree slice of phase choice. Even transposing inputs would only bring that to 320 degrees. Not only that, dynamic range is limited by a very poor choice of operating parameters for the FET driving the phasing system.

To have a perfect product you need 360-degree rotation with no level change. Even with perfect design, phasing systems are tedious to adjust. Other systems rotate phase over a wide range, but have severe level change as phase is rotated. The effect of a level change is you have to "walk" back and forth with phase an level every time you adjust phase! When level changes with phase adjustment, phasing units are even more unpleasant to adjust than a proper system is.

With a ****well designed**** bridge system, amplitude shift is minimal. The MFJ1025 series, despite being manufactured as cheaply as possible, outperforms any other unit I have tested (by a large margin) over its intended frequency range of 1.8-20 MHz.

The phasing system at W8JI, used to pinpoint signal directions of transmitters or jammers more than to actually cancel noise, is a more sophisticated version of the 1025 with higher dynamic range. It uses tandem bridges, and is calibrated in degrees shift. Phasing is spread over a wide, linear, control range. Phase rotates a full 360-degrees with less than 1dB amplitude shift. It is possible to resolve as little as 1- degree directional difference in arriving signals.

A Word About Dynamic Range and Noise Floor

160-meter DX'ers have the most demanding noise floor and dynamic range requirements of anyone

using a phasing unit. 160 ops not only must hear exceptionally weak DX stations using small inefficient antennas and low power, they must contend with multiple 1500-watt signals (often with excellent antennas) parked just hundreds of Hz away! CW operators often listen to CW signals with 250Hz BW systems, pushing noise floors up to 50 times lower than 10kHz filters would produce at any location. Amplifiers and other components must also handle the large powerful window of AM BC stations just below 160, as well as the SW BC stations above. At my house the accumulation of signals from stations *outside* Amateur bands is enough to light a small 12v light bulb when using a 15dB amplifier!

Obviously what works well at MW for sorting 10kHz spaced wide bandwidth signals apart is entirely different. Here we have the strong window of stations, but we are looking in that strong window for other stations that often are strong (compared to 100-watt transmitters with poor antennas). Noise floor is also much higher, and so dynamic range and noise floor are of almost no concern. The critical performance parameter is mostly adjustment range, and not dynamics.

SW BC listeners need something a bit different yet.

Everyone should be cautious assuming a device optimized for one application somehow makes it fit the other.

The only drawback of the higher noise floor of the MFJ-1025/1026 is you ***must place it after any amplifiers used with low-level low-noise antennas.*** The MFJ-1025/1026 noise floor is typically around 14dB.

The dynamic range and noise figure of my personal unit is excellent. Unfortunately such a system is impractical for amateurs and SWL's because of cost. Each unit has about $300 in raw material cost, plus the assembly time. The FET's alone are 28-volt devices capable of several watts RF power while providing a 1.5dB noise figure. But for typical operation the 1025 is almost as good. If you want to spend several hundred dollars, I can sell you a system that works almost perfect in almost any application!

Even with MFJ's well-known QC issues and affinity for building things cheap (not good) I still recommend the MFJ-1025 over all other mainstream units. It has the potential to be better than anything else with only minor changes, and is certainly betetr than all other mainstream units like the ANC-4 or QRN Squasher just as it is shipped. If you want a custom modified unit, and can not make the changes yourself, I would be happy to help.

GAIN MEASUREMENTS

The following table are gain measurements for the MFJ-1025/1026. Wiltron Network Analyzer (50 ohm detectors), transmission loss mode.

Stock Unit and Filters 50-ohm input and output

Aux Port Gain

Frequency	Gain (dB)
0.5MHz	-35
1.2 MHz	-16
1.5 MHz	-9.3
1.7MHz	-6.7
1.8MHz	-5.4
1.9MHz	-5.2
2.0MHz	-4.8
2.5MHz and higher	-4.4

From the above losses, deduct -2dB *(or add 2dB!)* ***if the input protection light bulb is bypassed or replaced with a jumper wire. I.e. 1.8 MHz gain becomes -3.4dB with the lamp removed.***

Main Port Gain

Frequency	Gain (dB)
0.5MHz	-34
1.2 MHz	-14.1
1.5 MHz	-8.0
1.7MHz	-5.8
1.8MHz	-5.0
1.9MHz	-4.2
2.0MHz	-3.7
2.5MHz and higher	-2.3

Bypass is essentially lossless.

Gain Compression and IM Dynamic Range

With gain controls wide open and pre-amp disabled, blocking begins at about -3dBm. Third order intercept is about -21dBm. This is generally well beyond what most receivers are capable of obtaining. (Click here to see the **receivers** page on this site.) The drawback of the 1025/1026 is the lack of any form of filtering, other than the high pass filter. This means the MFJ-1015/1026 sees a very wide window of signals.

With that in mind, it is questionable if removing diodes actually is worth the risk of RF or ESD damage. It would be much better to add a simple bandpass filter system for the band you are operating. However, there is some improvement offered by removing diodes. IM and blocking also can be improved by increasing quiescent current in Q5 and 6, as well as Q4. The bias modification is generally too complex for most people to do, and adding a bandpass filter in from of the unit would be much more effective anyway.

Removing Diodes and Lamps

IM dynamic range will improve about 6dB with this modification. The lamp will improve gain and noise figure of the AUX port (reduce loss) by about 2dB.

If you *never* transmit through the unit:

1.) Remove C5

2.) Remove D1,D5 at the output, remove D12 and 13 (or D10 and 11 in 1025) and D6 and 9 at the inputs.

3.) Replace Lamp1 with a jumper wire.

NOTE: If you transmit through the unit do NOT remove C5 and D1 and D5.

Sensitivity to Load Impedance

Like any typical amplifier system, this unit is sensitive to load impedance. Since it has an emitter follower at the output, it likes to see a low load impedance. Gain will increase about 5dB if you add a 1:4 step up transformer at the emitter of Q4.

The following are modifications or changes to the MFJ-1025/1026:

Gain Improvement Mod.

Increases gain and dynamic range by approximately 5dB (assumes your receiver has a 25-ohm or higher input Z):

1.) Cut the trace from C6 to D1/D5 near the small dip relay.

2.) Add 1:4 step-up transformer, low impedance side towards C6 and high impedance side to load.

Transformers are available, including US shipping, for $6.

Filter Modifications

The MFJ-1025/1026 contains BCB filters that roll off at about 1.9MHz. To extend operation lower in frequency:

Remove L3, L4, L5, L6.

Short C8 and 16

Increase the value of C12 and 13 by the same factor as the decrease in frequency. For half the frequency, double C12 and 13.

Caution: DO NOT remove filters if you intend to use this unit for weak-signal reception on Amateur bands above 1.8 MHz, unless you are sure you do not have strong BC signals.

C12 and C13 mods will reduce upper frequency limit proportionally to the change.

Extending Phase Range

Phase range of the MFJ-1025/1026 is limited by the ratio of R16 to reactance of C12 or C13. Normally phase will rotate through at least 130-degrees. SW3B inverts phase 180-degrees, and this moves the phase shift to the other side of the circle. Typically the MFJ-1025/1026 has about 280-degrees of phase rotation.

The remaining 80-degrees can be covered a number of ways. You can swap inputs (only when using it entirely passive), or modify the phasing network. I also have successfully added a relay that swaps the outputs of the low-pass input filters. A second method is to add an additional capacitor in series with the ground lead of the bottom of R16 (the wiper remains grounded) back to the junction of C14 and R17.

I'd recommend swapping the inputs, it is easier and does not degrade amplitude response or bandwidth.

Receiver_Tests

[Home] [Up] [Sorted by IM3]

Here a few measurements of receivers, using 500Hz filters.

MDS is a measure of sensitivity. -135dBm is 10dB more sensitive than -125dBm. This number doesn't matter very much in what you actually hear. **The real test is if you hear a very noticeable noise increase when you connect an antenna to the receiver. If you hear an obvious noise increase when you connect an antenna instead of a dummy load, your receiver is sensitive enough! You should check sensitivity at the quietest time with the narrowest selectivity you use on every antenna you use.**

Contrary to folklore and hyperbole, there isn't a receiver sold today that can dig into noise more than others on CW based on sensitivity or the use of a DSP or multiple DSP systems. The exceptions are:

- Increased selectivity will reduce noise
- Poor AGC design or detector problems can cause mixing of signals and noise

If you read a review that claims a receiver made weak signals appear from nowhere, you better keep a wary eye on the rest of the review. It is possible for a receiver to be abnormally bad, but it is not possible for a receiver to work better than other properly working receivers based on sensitivity.

BDR is blocking dynamic range. This is the point where a strong signal either 2 or 10kHz just starts to make your receiver lose sensitivity. The bigger the number the better, ESPECIALLY at 2kHz spacing. **The number you want here is probably around 80dB or more if you live in a reasonably quiet location and work weak signals on crowded bands. If you run two transmitters on the same band or have a neighbor who operates near your frequency, you almost certainly need more dynamic range. I'm in a very quiet rural location and have very directive antennas, and 80dB blocking DR suits my requirements just fine most of the time.**

IMDR is intermodulation dynamic range. This is the single most important number when comparing receivers. This is where two or more strong close-frequency signals mix and generate a new phantom signal or multiple tones in a adjacent frequency SSB signal mix with themselves and make what sounds

like splatter. The measurement is made just at the point where the phantom signal level is high enough to interfere with the weakest signal your receiver can detect. IMDR is a measure of how badly your own receiver causes problems you might blame on other people. Bigger numbers mean better receivers. It is most important the **2kHz** number be good. The 10kHz test number doesn't mean nearly as much, because almost any radio is good enough at 10kHz or wider. **Some number above 80dB is enough to stay out of trouble 99% of the time. If you are in a noisy location, you obviously need less performance. 85dB keeps my receivers at the point where poor quality external signals cause nearly all off-frequency problems. With 85dB IM3DR only a few of the strongest stations cause my receiver to make its own internal problems.**

Be aware the very good Drake receivers in the test below are essentially gutted and rebuilt receivers. They DO NOT use Drake mixers or amplifiers! Stock R4C's suffer from severe problems. See the notes!

CLICK HERE TO SEE SORTED BY CLOSE-SPACED IM3 PERFORMANCE, THE MOST IMPORTANT TEST!

Model serial	MDS	2kHz BDR	10kHz BDR	2kHz IMDR	10kHz IMDR
FT1000MP 8N370165	-135	73.5	84	62* 75.5** note 1	70.5* 80** note1
FT1000D	-134	68	85	72	81
FT1000D/W8JI mods	-136	73	87	89	88
FT1000MK V stock	-131	74	86	60 note1A	77 note 1A

FT1000MK V W8JI mods	-131	80	89	85	90
IC-751A	-140	83.5	98	79	91
IC-756 PRO	-133	N/A	78	67	77
IC-7800	-127	80	98	78	88@5kHz
R4C S/N 17816 stock	-132	57	109	48	82 note2
R4C S/N 22291 stock	-139	80	116	68	86 note3
R4C 17816 med mod	-140	96	115	92	103 note4
R4C heavy mod	-143	127	131	118	119 note5
TS-870 Inrad 400Hz IF filter mod (AIP)	-139 (-125)	74 (82)	94 (98)	86 (81)	90 (91)
TenTec Orion	-129	98	100	92	92

Note 1: The FT1000MP has the same inherent problem as the FT1000 and FT1000D. A noise-blanker FET is left on *even when the noise blanker is turned off with front panel NB and NB-W switches*. This FET connects directly to the IF, and overloads when strong signals are present. Unlike the FT1000 and FT1000D, the front panel noise blanker level control affects the FT1000MP IMD performance. * is with the NB gain control fully advanced, ** is with the NB gain control fully off. ***In ALL cases the NB off-on switch is OFF!!***

Note 1A: Yaesu copied the same basic NB from the FT1000D, and so this radio also has a NB problem even when the NB is switched off. This test was with NB gain menu set for minimum gain. The radio is worse when NB gain is increased in the NB gain menu.

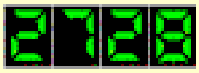
Note 2: This is a MOSFET 2nd mixer R4C, aligned and in good shape but with no changes. This receiver is almost useless for weak signal CW work of any kind because of poor IM performance and filter blow-by.

Note 3: This is a tube 2nd mixer R4C, aligned and in good shape but with no changes. This receiver is almost useless for contesting or crowded bands because of poor IM performance and bad filter blow-by.

Note 4: This is the same receiver in note 2, but with the following changes:

1. Increased screen voltage on RF amplifier to 130 volts.
2. 6J6 first mixer with oscillator injection increased to 5 volts.
3. Diode double-balanced 2nd mixer with +10dBm injection.
4. 10dB of gain removed from first mixer and first IF and redistributed behind the narrow filter.
5. Sherwood Engineering 600Hz roofing filter.
6. New AGC system. After narrow filter only, no AGC on front end.

Note 5: Gutted and rebuilt with solid-state double balanced high-level mixers. No vacuum tubes in signal path. Sherwood 600Hz roofing filter.

This page has been viewed  **times since revised on June 28, 2004**

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Receivers

[Home] [Receiver_Tests] [FT-1000 Noise Blanker Mod] [Orion and IC-7800]

Related pages: Checking signal BW FT1000 NB FT1000MK V NB External receiver preamps

My receiver Tests

Sherwood's Tests

We hear many things about R4C's (and other receivers), we love to talk about radios. Unfortunately most information comes from hearsay, useless subjective opinions, or from subjective reviews.

The only commonly published sources of measurements are ARRL or RSGB reviews. While the ARRL and RSGB do a good job of reviewing equipment, they publish somewhat useless wide-spaced data for receivers. Even with excellent wide-spaced performance, close spaced performance can be horrible!

Why Test at Wide Spacing?

Most manufacturers and many magazines test at wide spacing. When we have on-the-air interference problems, it is almost always with a station a few kHz or less away. Why would most of us care about a test or data at 20kHz or wider, when the bothersome signals are a few kHz up or down from us? Wide-spaced tests inflate performance, and gives us meaningless numbers for real-world performance.

Wide-spaced testing only evaluates RF amplifier and first mixer performance. Common design problems are easily and often missed when wide spacing is used. The weakest link is almost always downstream of the first mixer. There are several specific examples I can give:

1.) Receivers using DSP-based filtering systems for primary narrow selectivity.

2.) Receivers with poor 2nd mixer design (the R4C)

3.) Design errors in noise blankers, such as the **Yaesu's noise amplifier** design error

It's no wonder receivers have shown very little performance improvement over the years. Manufacturers evaluate performance on nearly useless wide-spaced measurements. They obviously only want to pass the wider test signal frequency spacing tests, because that is what we look at. Looking at 10kHz or wider tests, we all assume things are getting better. In actual use, most of our problems come from signals nearly on the same frequency, not 10kHz and especially not 50kHz away!

The only truly valid performance test is one where BOTH test signals are within the roofing-filter bandwidth. When close-spaced performance is good, wide-spaced performance is just as good or better. This is true for older radios and modern radios.

We also need to be factual in performance assessments. Too many feelings get in the way of being objective.

Receiver Myths

PIN Diode Mods

Save your money. PIN diode replacement never has changed performance in any receiver I've listened to or measured. PIN diode mods don't change distortion, blocking, noise, or any other parameter. A normal signal diode with proper bias is just as good.

PIN diodes function as "RF switches" or "linear RF resistors" only when the carrier lifetime exceeds the period of an RF cycle by a large margin. Most of the diodes used in PIN mods do not have long enough carrier lifetime to even behave like a low-distortion linear resistance, let along a pure switch. They really aren't any better than manufacturers stock diodes, with the PIN diodes barely being linear at 30MHz let alone at 455kHz. The whole "PIN diode thing" would be laughable if it wasn't costing people money!

If you have even measured a real difference, please e-mail and tell me what radio it was and what the test conditions were. If you are only going by emotion or feeling, don't bother reporting that. If I spent several hours and/or a few hundred dollars changing diodes and could not A-B the change, I'd probably think things got better also.

Drake R4C

The R4C is elevated to a status far above the realities of its actual performance. The R4C, like many DSP radios, has a wide performance variation between close and wide spaced tests. This is

especially true with the early S/N MOSFET mixer R4C's. With a wide-test, all R4C's look "good". That's because a wide test only checks the tube-type RF-amplifier and vacuum tube 1st mixer. Even though the R4C RF amp does better with higher screen voltage and the first mixer suffers from very low injection, the first two stages are still reasonable.

The problem is in the Drake 2nd mixer. The second mixer in the Drake ranges from poor to useless. The MOSFET mixer versions of the R4C are identifiable by looking at the MODE switch. If the MODE switch has CW 1.5, .5, and .25 positions the receiver has a tube (6BE6) 2nd mixer. If it is labeled CW1 and CW2, it has the horrible MOSFET mixer.

A few Drake website claims there is little or no difference between early serial number R4C's using MOSFET 2nd mixers and later R4C's using vacuum tube 2nd mixers. Nothing could be further from the truth. Worse yet, these pages steer people into a wide 6- or 8-kHz roofing filter for weak signal CW work, a very foolish choice as we will see by actual measurements. For serious CW work, the Sherwood Engineering 600-Hz roofing filter is an absolute must! R4C's have far too much filter leakage to be useful on crowded bands using CW with a wide roofing filter.

Roofing Filters

In order to significantly improve close-spaced performance a roofing filter has to attenuate signals on ***adjacent*** channels. We could have strong and weak signals CW alternating every 500Hz across the dial, assuming our transmitters were cleaned up. In the case of CW a roofing filter, to be useful, needs to be about 500Hz wide or less.

For SSB, we could consider channel width 3kHz. A roofing filer really needs to eliminate the adjacent channel up and/or down, and that means a roofing filter that passes only 3kHz or less.

While the typical wider roofing filters do a good job knocking down problems caused by signals up or down the band several kHz, they don't do anything for signals within the bandwidth of the roofing filter.

Testing Radios

TEST SETUP

My test setup uses two low-noise crystal oscillators. One oscillator is fixed on 1840-kHz, the other oscillator is selectable at 1840.5, 1842, or 1850-kHz (***note***: I now use variable frequency generators). Both oscillators use low-noise CATV transistors, and provide 20dBm output. These oscillators each feed a 1dB per step attenuator, with a total attenuation of 160dB available.

The attenuated outputs are fed to a "Magic T" combiner. "Magic T's", like every low-loss passive combiner system, are load impedance sensitive. Any mismatch reduces generator port-to-port isolation. In many cases, marginal generator-to-generator isolation can cause IMD in the signal sources. This generator-sourced IMD corrupts readings, changing IMD performance. To reduce combiner mismatch, the output of the magic "T" feeds a small ~3dB attenuator.

Signal level from each individual source can be varied in 1dB steps from +20dBm to -140dBm. +13.5dBm is the maximum level available, after combiner system losses of 6dB are added, for a final receiver signal range of +13.5 to -146.5 dBm. This range is ideal because my own transmitter is typically +15dBm on my closest receive antennas, while my daytime 250Hz bandwidth noise floor is near -140dB. Most of my receivers have sensitivities in the -140dBm range, allowing them to marginally get down to noise floor in the daytime. (Noise increases 10-20dB at night, on quiet nights, because of distant noise sources that propagate via sky-wave.)

Note: *Most setups use a single attenuator after the combiner, I chose not to do that. To reduce generator IMD, I decided to attenuate each signal source with carefully matched attenuators. If a test requires 30dB of attenuation, generator-to-generator isolation will be the sum "magic T" isolation and each attenuator pad isolation. In this case, generator-to-generator isolation would be well over 90dB, far more than I could obtain with the "Magic T" alone.*

The Magic-T combiner also has a 3dB 50-ohm output attenuator. This pad helps stabilize load impedances seen by the "Magic T", insuring return loss is at least 6dB. Total attenuation through the "Magic T", including the internal attenuator pad, is 6.5dB. The port-to-port cross talk of the magic "T" can be nulled with a small trim pot. This adjustment is only necessary when doing tests with near-zero attenuation and a mismatched receiver.

MEASUREMENTS

Since virtually every receiver overloads in stages after the attenuator or pre-amplifier, there probably is no compelling reason to measure receivers with the attenuator on. Adding an attenuator will not increase the dynamic range, it will simply move the raw measurement numbers higher. In other words, we don't care what the absolute numbers are...we can always add or remove gain external to the receiver. What we do care about is *dynamic range* of the system, both for blocking (a strong signal makes the weak signal disappear or get noisy) and intermodulation products. The lower the dynamic range numbers, the worse the receiver will be.

Measurements involve three basic procedures, all measured in dBm (dB milliwatts).

MEASURING MDS

A conventional signal generator is used to measure minimum discernable signal (MDS). This point is where the signal is just clearly audible, about 3dB out of the noise floor.

MEASURING BDR

Blocking Dynamic Range is measured by setting one oscillator (or the signal generator) to a test frequency either 2 or 10 kHz above the interfering signal. This test is equivalent to having a single strong station come on a certain amount away from a very weak station you are trying to copy, and making your receiver lose volume or have a hiss that increases compared to the weak signal's level.

To minimize generator noise, the low-noise crystal oscillators are used for the strong signal. This creates a "perfect" zero bandwidth strong signal with virtually no broadband noise.

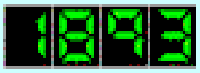
The level of the strong signal is adjusted until the slightest detectable change in S/N ratio occurs. The difference between the MDS and the level causing the blocking is the blocking dynamic range. This will be the ratio of the weakest to strongest signal the receiver can handle without losing a noticeable amount of weak-signal sensitivity, if the strong signal is a perfect signal and the weak signal is right at noise-floor.

MEASURING IMDR

Intermodulation dynamic range, or two-tone dynamic range, is measured by running two equal strength signals (from the low-noise oscillators) into the receiver with a certain test spacing. This test is equivalent to having two strong signals very near each other, with just the right spacing to cause a mixing product to fall on top of a noise-floor signal you are trying to copy. When the signal level of the mixing product is just audible above the noise floor, the ratio of the strong signals to the MDS (minimum discernable signal) becomes the IM dynamic range.

Poor IMDR performance shows up as splatter on SSB and as bloops, bleeps, and random musical thumps or phantom signals on CW.

Look at some receiver measurements!

This page has been viewed  since rewritten on June 28, 2004!

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FT-1000 Noise Blanker Mod

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This mod is part of a plug-in package available from me. See the Keyclick mod page.

I've tested over 15 FT-1000 and 1000D's with this mod. This modification will reduce close-signal IM in every FT1000 or 1000D by a substantial amount!

The IM problem is created because the FT1000 heavily forward-biases the noise-blanker's first FET whenever the blanker is turned ***OFF.*** This causes the FET hanging on the IF system to have high gain. Strong signals within roofing filter BW saturate the FET's drain, causing mixing products (IM).

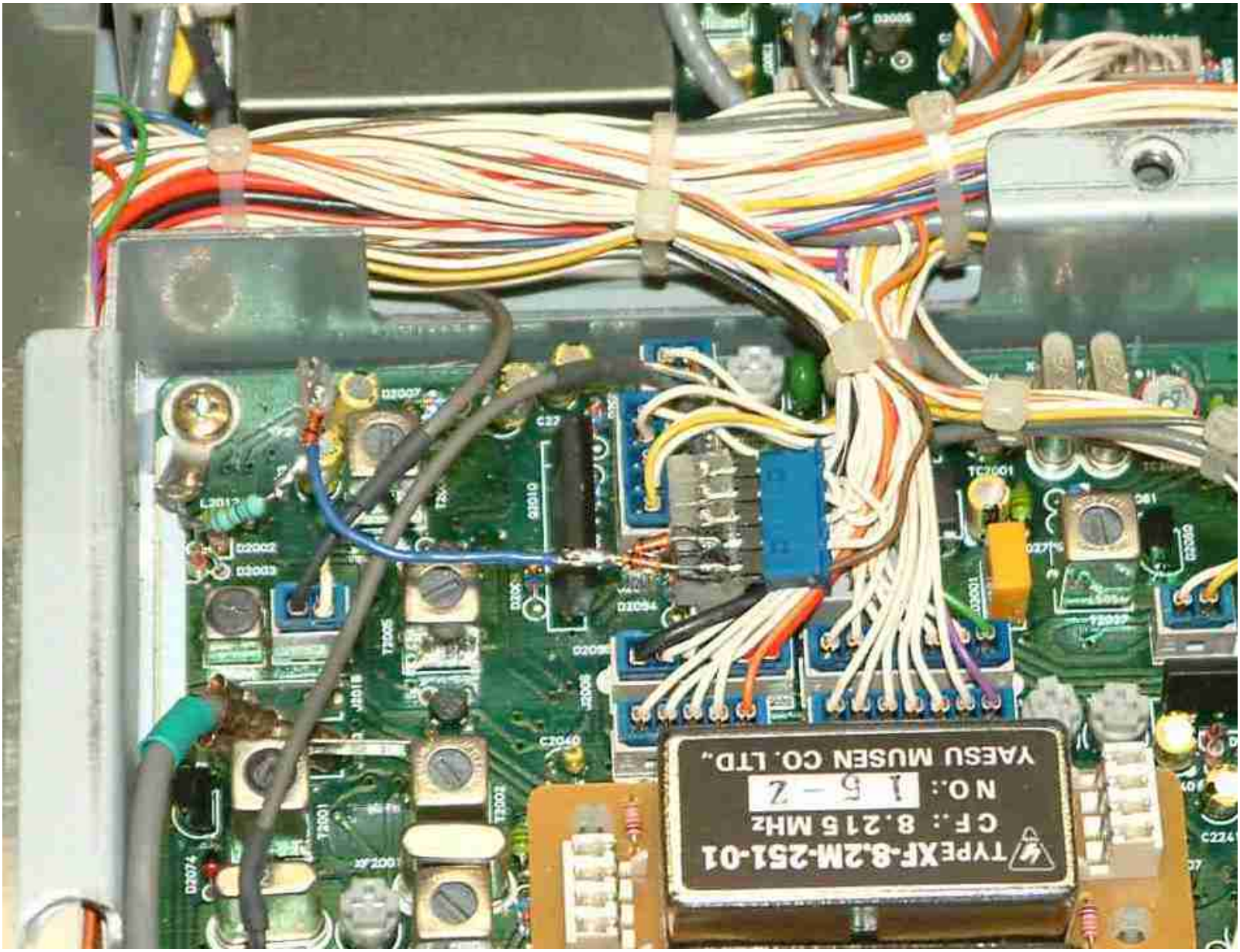
These new signals feed directly back into the 8MHz IF section, creating interference to any desired weak signals. The FET's IM creates new "phantom" signals that sound like normal CW signals. On SSB, it creates IM that sounds exactly like splatter!

This modification removes harmful forward FET bias when the noise blanker is off, and prevents the noise-blanker from creating interference to weak signals when the blanker is turned off. The NB functions normally when on.

The Mod

1.) Remove the bottom cover.

2.) Locate the noise blanker on the left rear corner of the IF board near the first group of 8 MHz filters.



Download Picture

3.) Locate TP-2001.

4.) Connect a small silicon switching diode (1N4148 is one example, although other silicon-type signal or switching diodes will work) through a 10k ohm resistor to ground from TP-2001. The diode cathode (banded end) goes towards the resistor and ground, anode towards TP-2001.

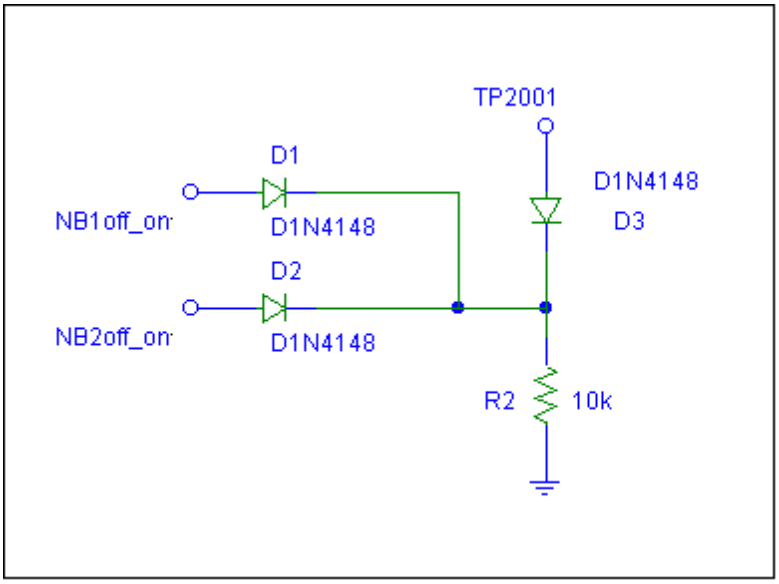
Caution: DO NOT use a high leakage germanium diodes. Use low-current silicon switching or detector diodes.

5.) Construct a pair of diodes with the cathode (banded end) tied together. Leave the leads long for now.

6.) Connect one of this diode pair's anode to pin 1 (the brown wire) of J2001.

7.) Connect the other diode anode to pin 2 of J2001.

8.) Connect the common-cathode of these two diodes to the junction of the 10K resistor and the cathode of the diode installed in step 4.



Note: I just pushed the wires down in the pin of the plug (P2001) and tacked them in place with a tiny bit of solder. You can verify these are the correct pins by measuring the voltage when switching the NB and wide NB buttons. Turning either or both NB switches "on" applies about 8 volts to either or both pins. Connect the common to the diode cathode and 10k resistor junction through a short hook-up wire.

This mod removes forward AGC bias from Q2003 and 2004 when the noise blanker is off, and prevents IM distortion fed back via C2041 and foil traces from impacting the receiver when the noise blanker is off. Other than improving IM performance when the blanker is off, it has no effect on NB performance.

Keyclicks_1000D

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Revised 1/31/04 at 0130Z

Receiver Improvement for FT1000 and 1000D

The FT-1000(D) runs an essentially unshaped CW waveform into the filters. This harsh signal generates unnecessary key clicks. The clicks are strongest from almost 1 kHz below to 2kHz above the transmit frequency.

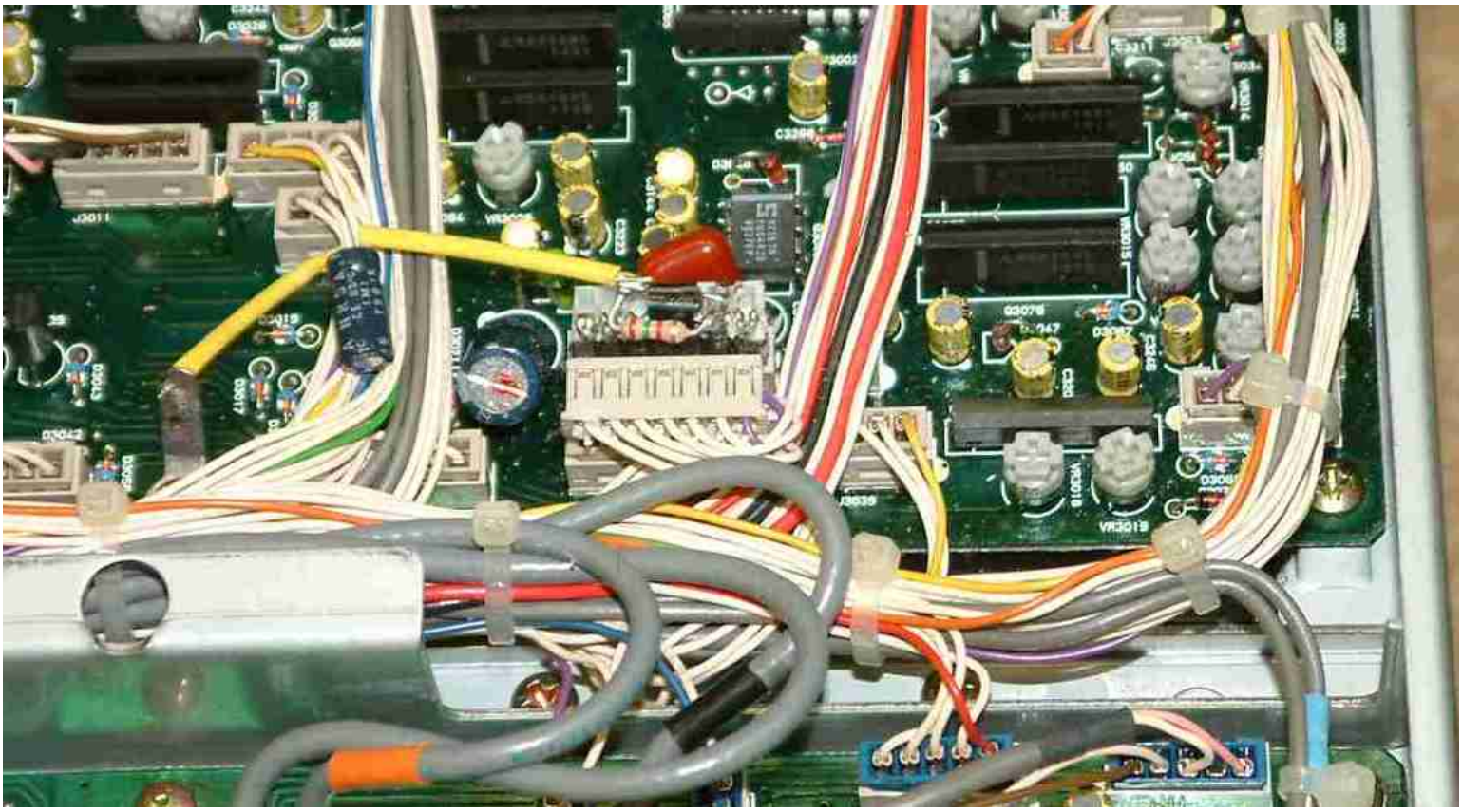
The cure:

I now have a plug-and-play 100% tested mod for the FT1000 and 1000D. This mod just plugs in, and includes the receiver NB mod.

Neither mod requires soldering! Only a screwdriver is required. The price for both mods is $45.

View plug in mods.

Locate the seven-pin connector J3024 at the middle-front of the AF board. (The AF board is the one with the audio, PTT, ALC, and key jacks as well as other connectors.)



This is a picture with a hand-wired mod in place.

Download picture

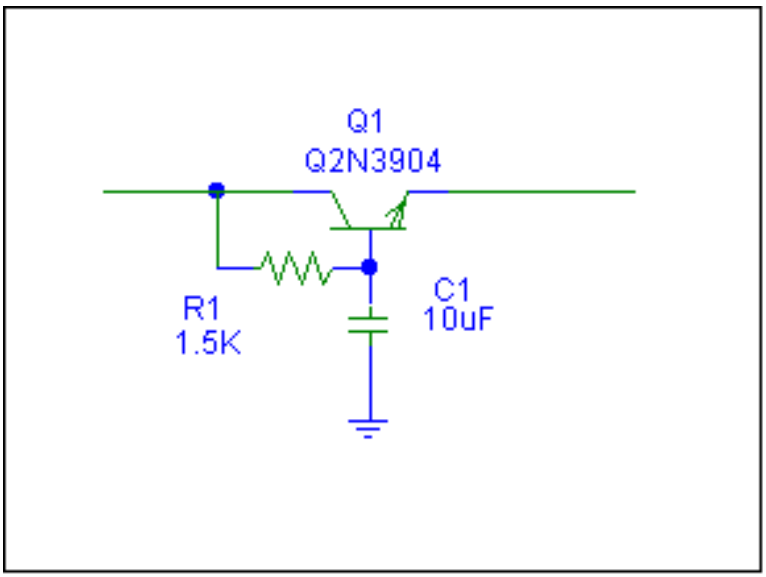
Cut the wire going to pin 2 T CNTL). This is the second terminal from the right (front of radio viewing). This is NOT the lead with 9 volts (nine volts is on pin 6). If you are unsure you counted the correct direction, measure the voltage!

Connect the collector of a 2N3904 to the flying harness wire you just cut.

Connect the emitter towards the jack on the wire you just cut. This transistor MUST be a 2N3904.

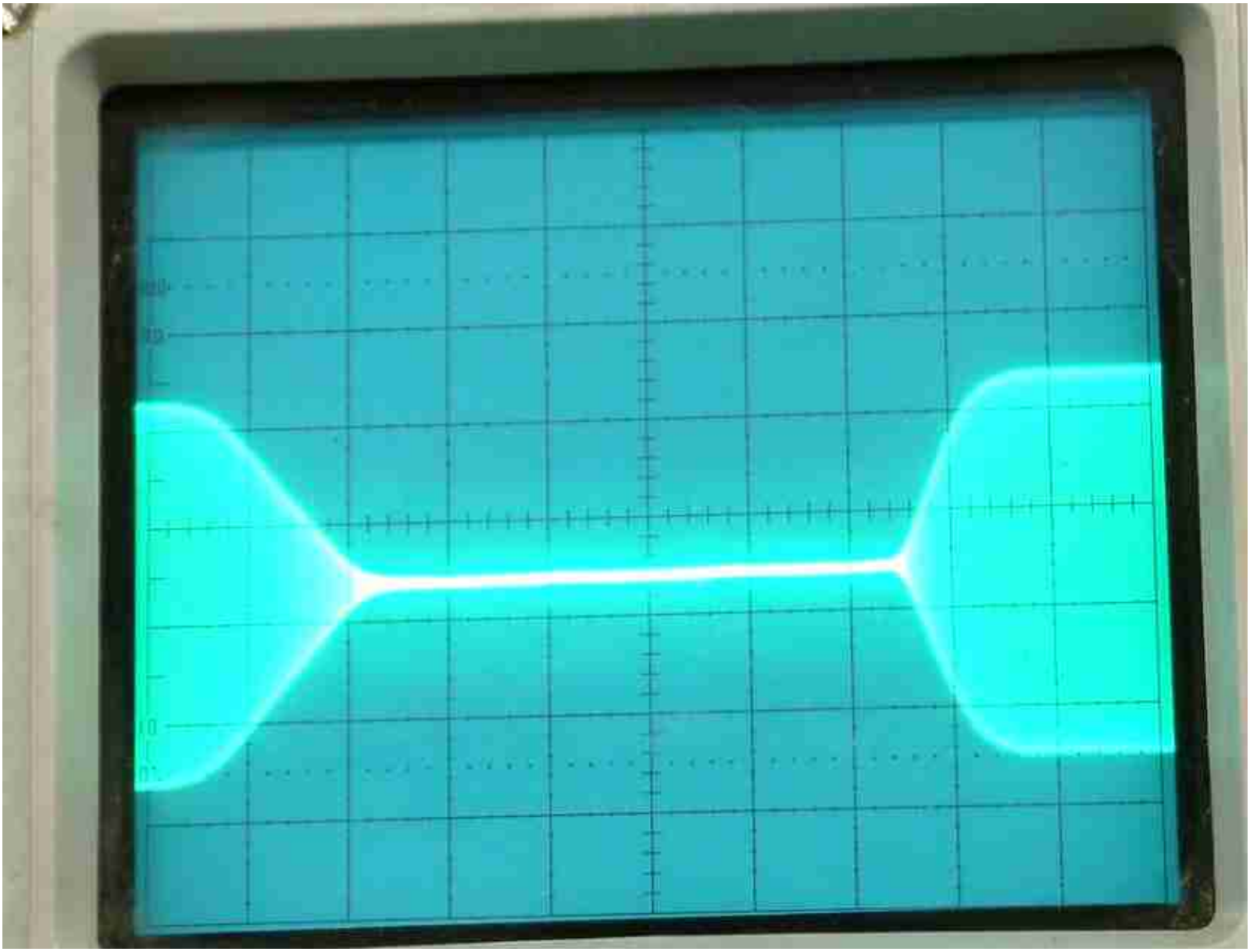
Connect a 1.5k resistor from collector-to-base on the transistor.

Connect a 10uF, 25-50 volt capacitor from base to ground. A non-polarized capacitor may be better, if you can find one. No one has reported problems with normal electrolytics as long as the positive lead connects to the base of the transistor.



The Results:

This modification slows the waveform rise and fall times at Q3030 and D3009. The result is a normally sloped CW waveshape into the SSB filter.



This is a nearly-perfect waveform with about 7mS fall and 4ms rise. The important thing is there are no sharp edges that cause clicks!

Clicks beyond 1 kHz above and below the transmitting frequency move into the composite transmitter noise, and there is about 50 dB reduction in clicks on some radios above the TX frequency!

Clicks 500Hz away are reduced about 20-35dB.

There is almost no change in "keying sound" on the transmitting frequency. The envelope shape is near ideal, and bandwidth reduction is dramatic.

Operating

Warning: This mod is NOT compatible with QSK with the full 10uF in circuit. For QSK use, I suggest adding a SPST switch from the 10uF capacitor to ground. You can place an additional 10uF capacitor across that switch, or simply add an additional 10uF in series with the ground lead from C1 (this gives 5uF total value at the base of Q1) and *not* switch the system. When not operating QSK, I recommend using the full 10uF of C1 for the cleanest signal.

The ideal wave shape occurs when the ALC is at the upper area of the blue ALC line on the FT1000 meter. With full **DRIVE**, adjust for desired power with the **RF PWR** control. After achieving desired power, back the **DRIVE** down until the ALC indication just starts to move away from full blue area (half-scale). Anything from 1/3 to 1/2 scale is a normal waveform.

For lowest clicks, adjust **DRIVE** until ALC until it just registers (perhaps 1/8 to 1/4 scale on ALC).

Keyclicks

Bandwidth rules Part 97.307

Link to W9CF. Kevin analyzes key clicks mathematically.

I noticed W2VJN has proposed using absolute signal level at a fixed test spacing for click measurements. While I agree with George's suggestion (this link) we use an established test method, I strongly disagree that a 15Hz BW signal level measurement of the peaks accumulated in dozens or hundreds of scans is any more useful that listening off-frequency on a known good receiver.

While his mod is certainly better than a stock Yaesu, clicks also can be reduced more than his method provides without affecting ability to work weak signals. We don't need to watch output power and keep duty cycle to 50%, a slight adjustment of weight control would do the same thing.

The real point is that both George and my suggestions are patches, they are not cures. Yaesu should step up to the plate and be responsible, and set an example for others to follow.

Testing

We really need to establish a standard for measurement of transmitters on both SSB and CW, and it should be the effective spurious power level on close adjacent frequencies using typical CW bandwidth. Looking at one spot with slow sweep and peak storage on a spectrum analyzer is not a good test.

It's pretty easy to see using the level (at one specific frequency) of a long-time average of signal peaks with a narrow filter sweeping by (which is what spectrum analyzers do) is a bad idea.

Some radios have a sharp click on one edge, and nothing on the other. This causes peak energy to be higher in proportion to average energy. Radios can have higher peak level and actually do less damage to adjacent channels than a lower level click that hammers on both make and break, like the FT1000MK V does.

If I look on a peak sample and storage device like my spectrum analyzer, a radio could actually

look worse yet bother adjacent CW bandwidth channels LESS than a radio that appears to be better.

A very narrow filter swept over the frequency with average storage of peaks does not account for attenuation slope modulation sidebands. I'd rather have a radio next to me with a steep drop in clicks at 500Hz than one with a single-pole gradually shaped slope even if the gradual sloped radio had less level. A peak comparison at some specified spacing misses too much.

SSB vs. CW

CW shares the same testing problems as FM and SSB. The FCC has changed some commercial voice tests to focus on spurious power level on adjacent channels, rather than using conventional but somewhat useless two-tone tests. A two-tone test shows the very ***best*** a radio is likely to do, not the ***typical performance***. The FCC now requires normal modulation and an adjacent channel ***power level measurement*** in some cases.

CW and SSB testing for amateur use should be the same way.

I think is is shameful that manufacturers can't manage to spend a few hours of engineering time to select different component values, instead of sticking us with radios that have poor keying and IM characteristics. Frankly, there is NO excuse or justification for selecting wrong component values in a such a simple design area.

Is an excessively clicking radio legal?

Most of us agree right off the bat that some radios are annoying, and MOST people actually want to have the cleanest possible signal. Of course there are a few self-centered people who care less if they bother other people. They generally use the excuse that since a radio was sold it must be legal.

Here is a link that shows the exact 97.307 FCC text.

What do bad and good signals sound like?

For now, you can listen to some recordings and judge the difference.

Key clicks are off-frequency sidebands heard when there is no trace of a tone. If you listen carefully to these recordings, you will see the "hardness" of a properly filtered transmitter on frequency sounds very little different than the hardness of a wide, clicking signal.

Here are some sound files....these files are intended to be educational. While it isn't OUR fault the rigs were incorrectly designed, it is our legal obligation under 97.307 to correct the problems we have been stuck with. This is especially true when big antennas and amplifiers are used.

First, this is a mono recording of a virtually click-less radio. This station is 20dB-over-nine, and my noise is S2. When you listen, notice the "clicks" disappear right when any trace of tone vanishes. You'll notice the CW still sounds hard, and this station regularly broadcasts high-speed CW that hundreds of people copy. This is the ***Click-free signal*** of W1AW on 160 meters.

If you own a stock FT1000, FT1000D, FT1000MP or FT1000MP MKV (or any rig with a rise and fall faster than 4 or 5mS OR with any sharply rounded corners) you have work to do!

Notice key clicks can be heard long after the tone disappears totally, as we tune across the signal, yet there isn't any trace of receiver desense. S-9 plus 10dB station from England, 40-meters, using ***FT1000MP MKV*** tuning up 2 and down 2 using 500Hz RX filters. Significant clicks in the USA from England well after sunrise in England when signals are dropping. Imagine a 40 over MKV! This appears to be typical of all MK V's

S-9 +20dB in S-2 noise tuning from 1.5kHz below stock ***FT1000D*** (sorry, I snipped the above freq tuning to avoid call sign) This is typical of all FT1000(D)'s

40-meter signal S-9 plus 15dB, background noise S-2, 500Hz filters, tuning up and down 2, recording of ***FT1000MK V***. This rig cuts a 3kHz wide swath that is S-6 or stronger.

40-meter signal S-9 plus 25dB, S-1.5 noise. Tuning up 2 and down 2, 500Hz filters in RX. ***Omni-6*** (This might or might not be typical of all Omni-6's. Ten-Tec reports an internal adjustment, when incorrectly set, causes these clicks. I've logged several Omni-6's with this problem, and some users continue to receive complaints after trying suggested mods. The characteristic I seem notice is key clicks extend downwards much more than upwards in frequency.)

160-meter, S-9 plus 15 dB signal, S-1 noise, tuning up and down 2, Inrad-mod in ***FT1000MP***

20-meter ***Kachina***. S-1 noise, signal S9 plus 10. Which signal is the "real signal"? A snip of a clean signal answering him, about the same signal level, is at the end. Spurs went down 2.4 and up 2.4, and were spaced 600Hz apart. (This might not be typical of all Kachina's)

40-meter signal, S-9 plus 10dB signal S-2 noise, tuning up and down 2kHz, 500Hz filters, stock ***FT1000MP*** (this is typical of all stock FT1000MP's. If you own one that has not been correctly modified, you can be sure it has grossly excessive key clicks)

160-meter signal, S-9 plus 20dB S-1 noise, tandem 250Hz filters in very high-dynamic range CW-only receiver, tuning up and down 1.2 kHz, stock ***IC765***

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FT1000MK_V

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Preliminary tests show very strong keyclicks +1kHz and -1kHz. The LSB clicks peak 500Hz below the TX frequency, and the USB clicks (in the LSB mode) are very obnoxious because they are at twice the rate as LSB clicks (strong on both make and break).

The FT1000MK V shows a 1mS rise and a 2ms fall with sharp edges and a very poor rise and fall slope.

Keyclick and Noise Blanker Mod for FT1000MP MK V

It's great to see a manufacturer offer improved close-spaced SSB transmit performance. The MK V Yaesu reverses other modern radio's downward spiral of transmitter SSB IM performance. Yaesu included a class-A mode.

Unfortunately close-spaced receiver performance has been neglected, and this affects all modes. The transmitter also has nasty CW clicks. Stock MK V's appear to be as bad as or worse than older 1000 series radios in two important areas:

- Clicks appear to be more severe than those from the notoriously poor 1000MP's
- The noise blanker continues with the same circuit error as the FT1000/1000D/MP series

Since a portion of the keyclick mod requires accessing the same general area as the noise blanker, it would be prudent to fix the receiver and transmitter at the same time. This article offers a combined modification that patches both problems.

Keyclick Problem

Raised- sine rises and falls would provide the fastest possible CW speeds for a given bandwidth. With properly filtered rise and falls, we would hear little or no change or softness when listening on-frequency. Tuning off- frequency, clicks would quickly vanish.

If you want to hear the sound of proper shaping, listen to this click-free signal recording as I tune across the signal. Off- frequency (even a few hundred hertz), we hear no clicks at all. On- frequency the CW is "hard" sounding, allowing copy to 60-WPM or more. When the tone disappears in the deeper receive filter's skirts, clicks also disappear.

In contrast, compare the MK V recording as I tune past the signal. This signal is from Europe on 40 meters!

There is a day-and-night difference off frequency between the no-click and loud-click signals. On-frequency both signals sound the same.

MK V clicks are caused by excessively fast rise and fall, and very poor shape of the rise and fall.

Rumors sine-shaped waveforms impact tone or readability of signals are false, as are claims clicks rolling of at some "X-dB-per-octave rate" beyond a few hundred Hz are a necessary part of life. Such statements are misleading, likely being based on the incorrect assumption the receiver has very wide bandwidth and the transmitter is filtered through a single stage simple resistor-capacitor click filter.

Unfortunately when we patch poor CW transmitter designs, we can not make perfect corrections. Without major modification we can not modulate the MK V (or most other transceivers) with properly filtered (which also means perfectly shaped) rise and falls.

This modification, like the MP and 1000D click mods, is a patch...not a cure.

Patches Vs Cures

Because mods on existing radios are patches, the radio owner must make a choice. If the user operates speeds faster than 45 or 50 WPM, the rise and fall required for legal close- spaced operation may be less than ideal. This does not mean the ability to work weak signals at modest speeds (up to 30 WPM) would be compromised even the slightest amount. It means high- speed ops (speeds over 45-50 WPM) may find the CW slightly mushy when adequate for close-frequency operation.

If a 2 or 3 millisecond rise and fall is used for operation at very high CW speeds in a single-pole R/C filter (this radio uses a simple RC filter) a transmitter is almost guaranteed to interfere with less strong signals within 1kHz or so. Transmitters with fast rise and fall times should stay at least 1.5kHz away from operators working weak signals, especially when the CW transmitter has 1930-era transmitter CW shaping.

Operators with stock MP MK V's should always try to operate at least 4kHz away from weak or moderate signal- level stations. Part 97 rules prohibiting keyclick emissions that interfere with adjacent frequency operations. The specific rule is 97.307(b) "Emissions outside the necessary bandwidth must not cause splatter or keyclick interference to operations on adjacent frequencies."

The Click Mod

The actual click mod requires changing two stages. The first stage modified is on the IF board. The IF mod slows the rise and fall of mixer transistors Q2033 and Q2038.

Note: This stage is easy to modify, and is located on the same board as the noise blanker. This allows the noise blanker to be corrected at the same time.

By itself, modification of the easy-to-reach IF board is NOT effective for substantially reducing clicks. A later stage on the RF board also has truncated rise-and-fall times with a very poor R/C edge-shaping system. This later stage continues to add clicks even after earlier stages are modified.

RF amplifier stage Q1001 has the fastest rise and fall in the RF section. Q1001 must have proper gate bias shaping and timing to reduce clicks to acceptable levels. Removing D1002 and altering components around Q1002 slightly reduced clicking, but I concluded any effort wasn't worth the result with bias rise and fall more rounded at Q1002.

As designed, Yaesu uses a square wave very rich in harmonics to drive a simple R/C filter. This poorly filtered square wave amplitude modulates the RF and IF sections. The poor basic filtering design, combined with non-linear amplitude response, requires great care in component selection. It also means we never will achieve the optimum bandwidth for any give rise and fall time and ultimate CW speed.

The NB Problem

Signals inside the wide roofing filter BW of the MK V reach the gate of Q2009 through C2043. This point precedes the narrow 8MHz IF filtering, allowing a rather wide swath of unwanted signals to reach the gate of Q2009. Q2009 is left operating even when the noise blanker is turned off, and can have substantial gain depending on bias voltages at TP2001. Bias voltages at TP2001 can be varied by changing menu settings for NB gain, but never fully turn off Q2009.

The accumulated level of all signals reaching the gate of Q2009 produce a large net voltage at the drain of Q2009. This voltage (and resulting net current) causes overload and distortion in the non-linear characteristics of Q2009 and 2010 (my manual has very poor printing, but I believe it says 2010).

New signals created by this distortion and the resulting mixing products are fed back through C2043 to the IF strip. The IM products appear as "phantom splatter" on SSB and " phantom signals" on CW. We can not actually hear the distortion on frequency of strong signals. Instead the IM products appear in the form of artificial interference when we attempt to copy weaker signals within 10kHz of a mixture of signals containing a few moderately strong signals.

The NB Patch

The NB mod is a simple effective mod, and improves close-spaced IM3 dynamic range around 10dB. A simple factory change of moving one foil trace would have made the MK V receiver noticeably better, but fortunately this mod is fairly easy for owners.

The NB correction removes the surface mounted 220-ohm resistor (R2046) from the source of Q2009, replacing it with a 220 ohm resistor connected between Q2009's source at C2027 and Q2016's (2SC4047) collector and the junction of R2049 (220-ohm also).

Making the Mod

This mod is a little more complex than the FT1000MP mod because the chassis of the MK V is a little more complex and unfriendly. Like any service work, having a clean open bench and a spot to separately store screws and other hardware removed in every step in order will make the job smooth and easy. (If someone sends step-by-step text, I'll put it on this page.)

Some may wish to remove and change parts, but I prefer to wire the click-mod to a single terminal strip. This will allow you to customize the mod, switch the mod in and out, or correct any errors without dismantling the entire radio.

You'll need the following parts:

(1) one foot each of two small insulated wires, #20-#26 one (preferably) green and one blue to make connections

(1) four-lug (with ground) terminal strip

(3) .1uF 50 volt disc capacitors (C1-C3)

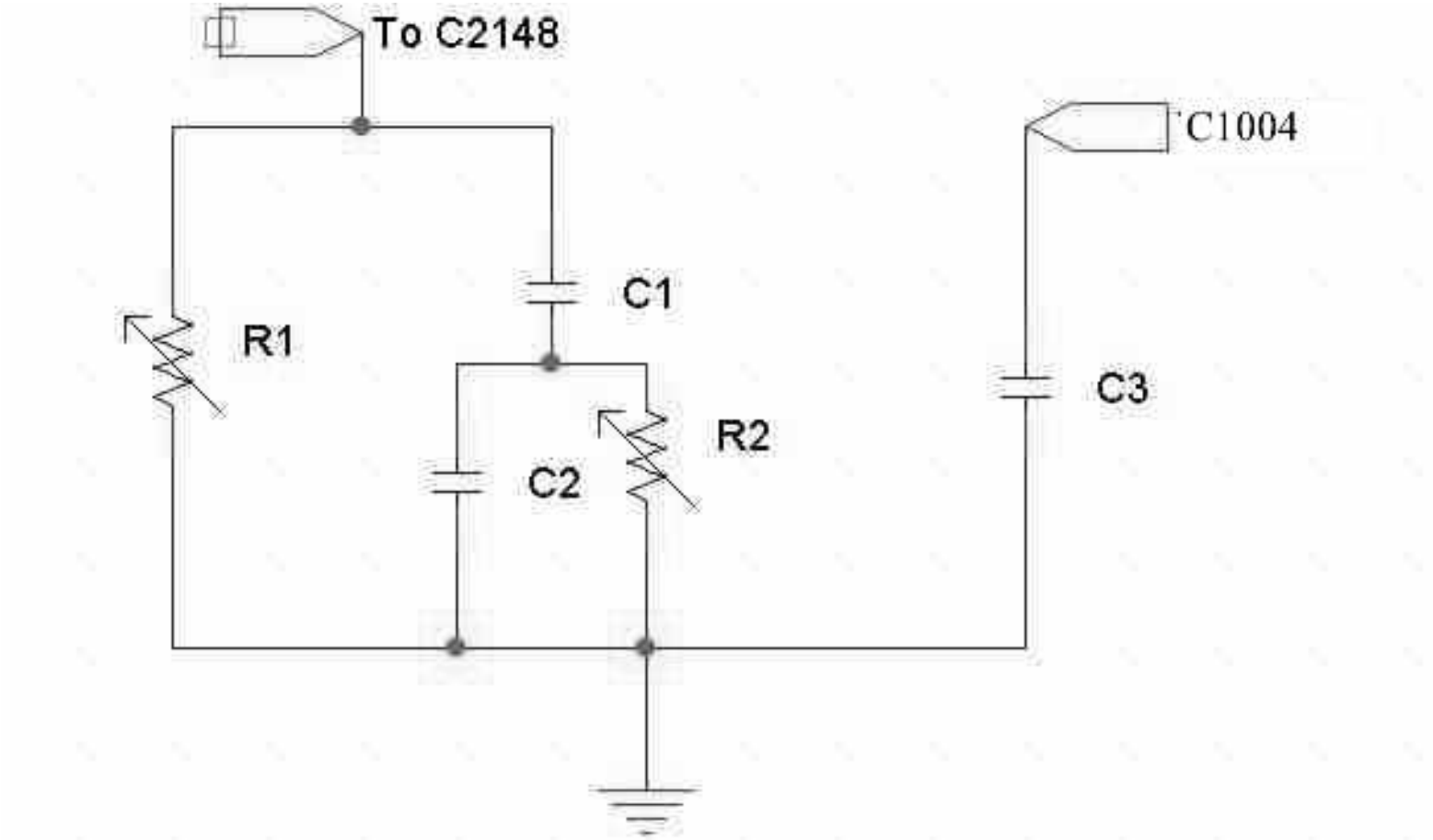
(1) 22k 1/4w fixed resistor (R2)

(1) 680k 1/4w fixed resistor (R1)

(1) 220-ohm 1/4w fixed resistor

In addition you need a well-lit bench, along with some hand-tools such as soldering pencil and solder, screwdrivers, and cutters and strippers.

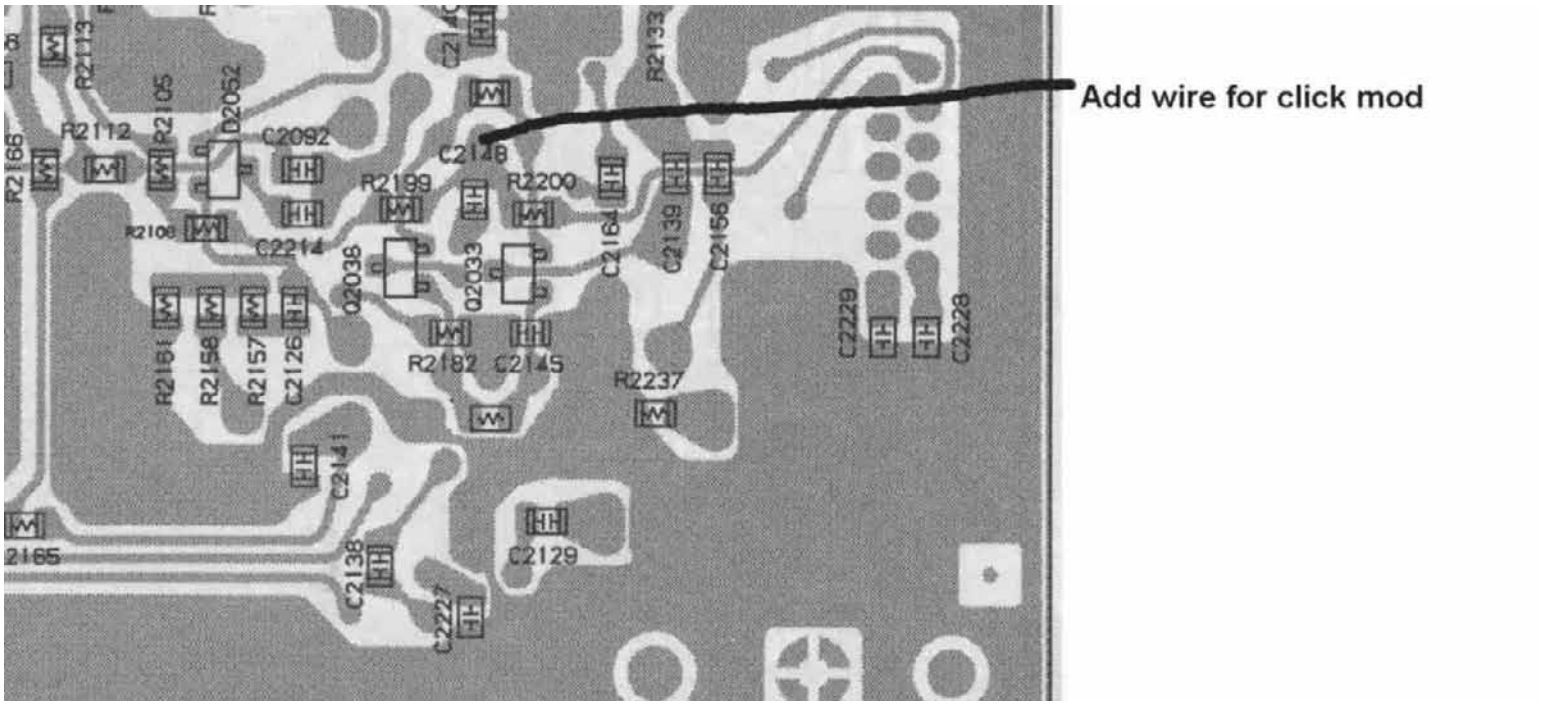
Populate the terminal strip as follows:



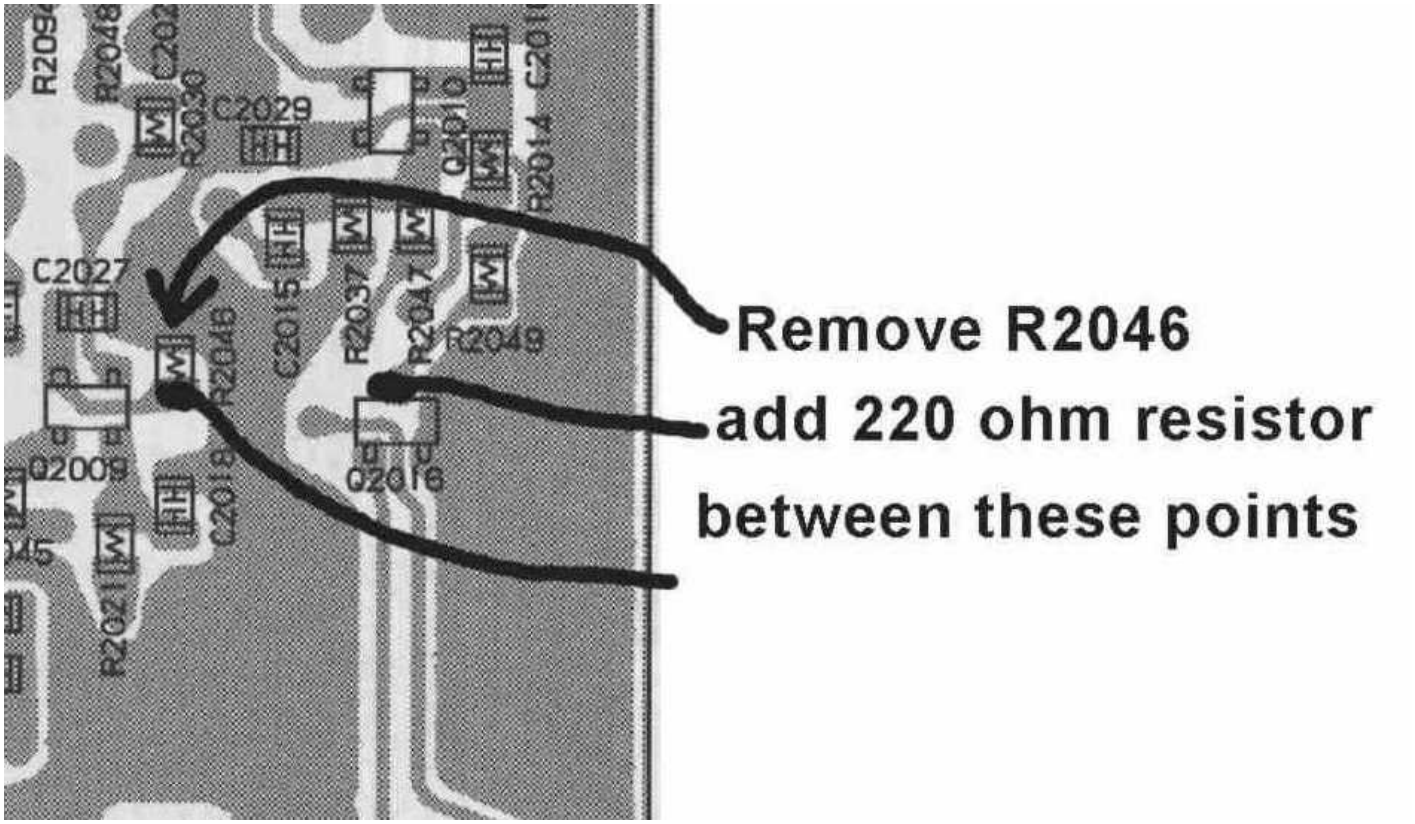
Remove top and bottom covers to gain access to internal circuitry

Remove the screws holding the IF board in place, and the minimal amount of plugs to allow flipping the IF board over. Draw a roadmap of all plugs that must be disconnected. This will help you remember where everything goes!

(click here to download expanded IF board if needed)



Attach one end of the blue wire to the ungrounded end of C2148, let the other end float



Locate and remove R2046.

Add the 220-ohm resistor to the Q2009/C2027 source and capacitor connection point

Connect the other lead of this resistor to the junction of Q2016's collector and R2048.

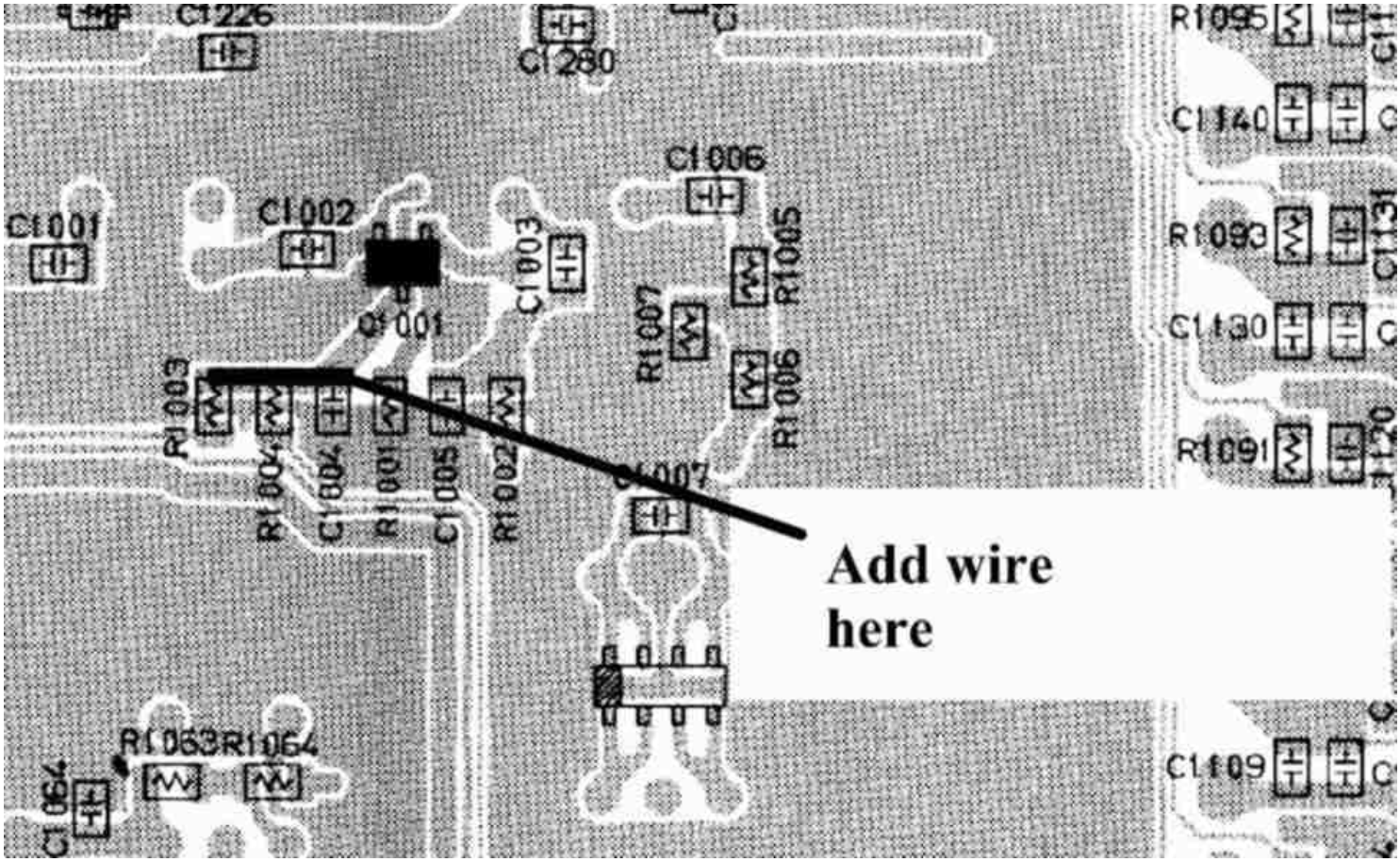
Reinstall the IF board with the flying lead exiting the closest edge of the PC board.

(Take care to avoid pinching any wires.)

Remove all hardware necessary to access the RF board

Remove the RF board

(Click here to download an expanded view of RF board if needed)



Attach the green wire to the junction of C1004 and R1003/R1004 at the gate of Q1001

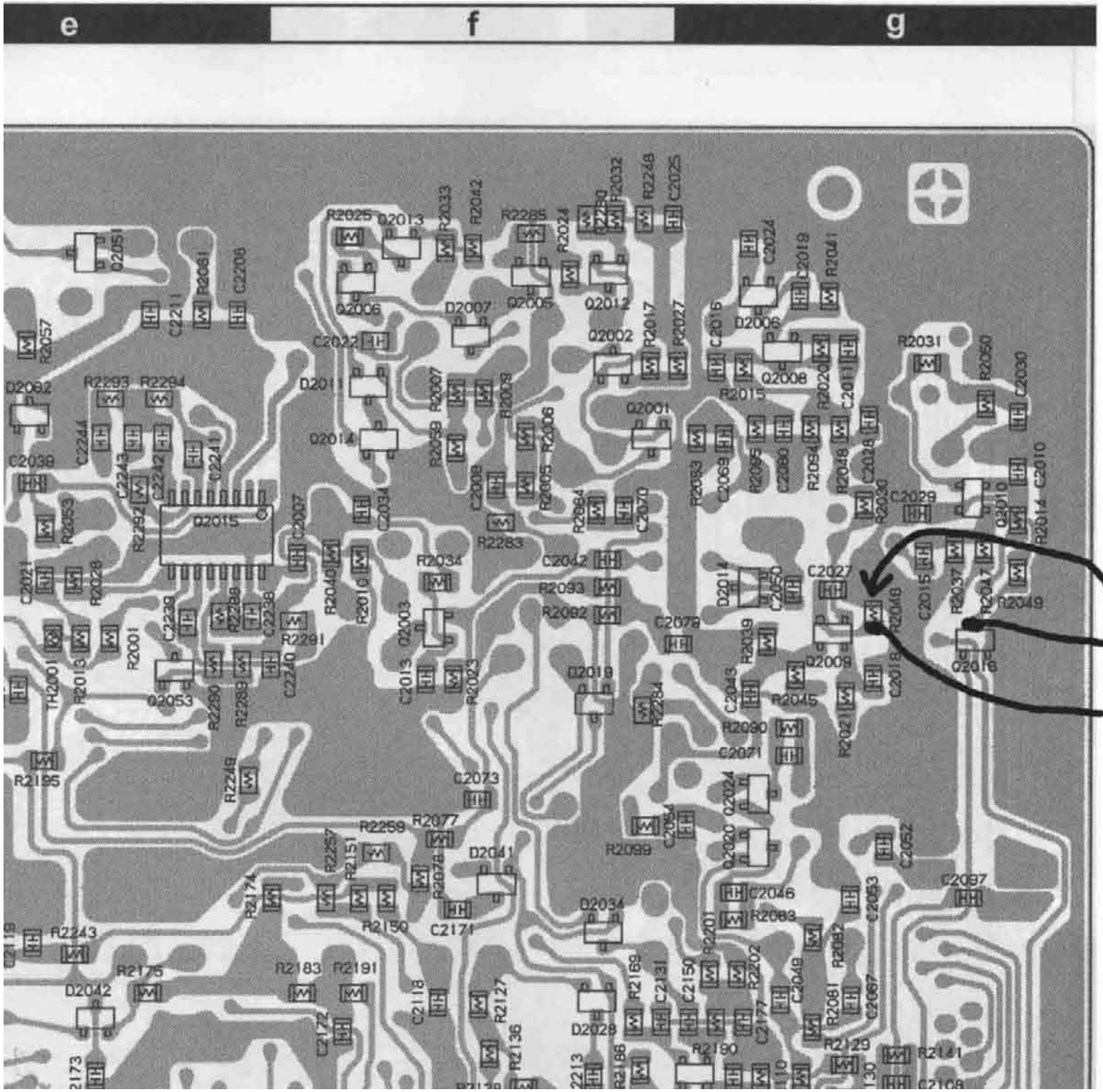
Route this wire up to the area of the IF board

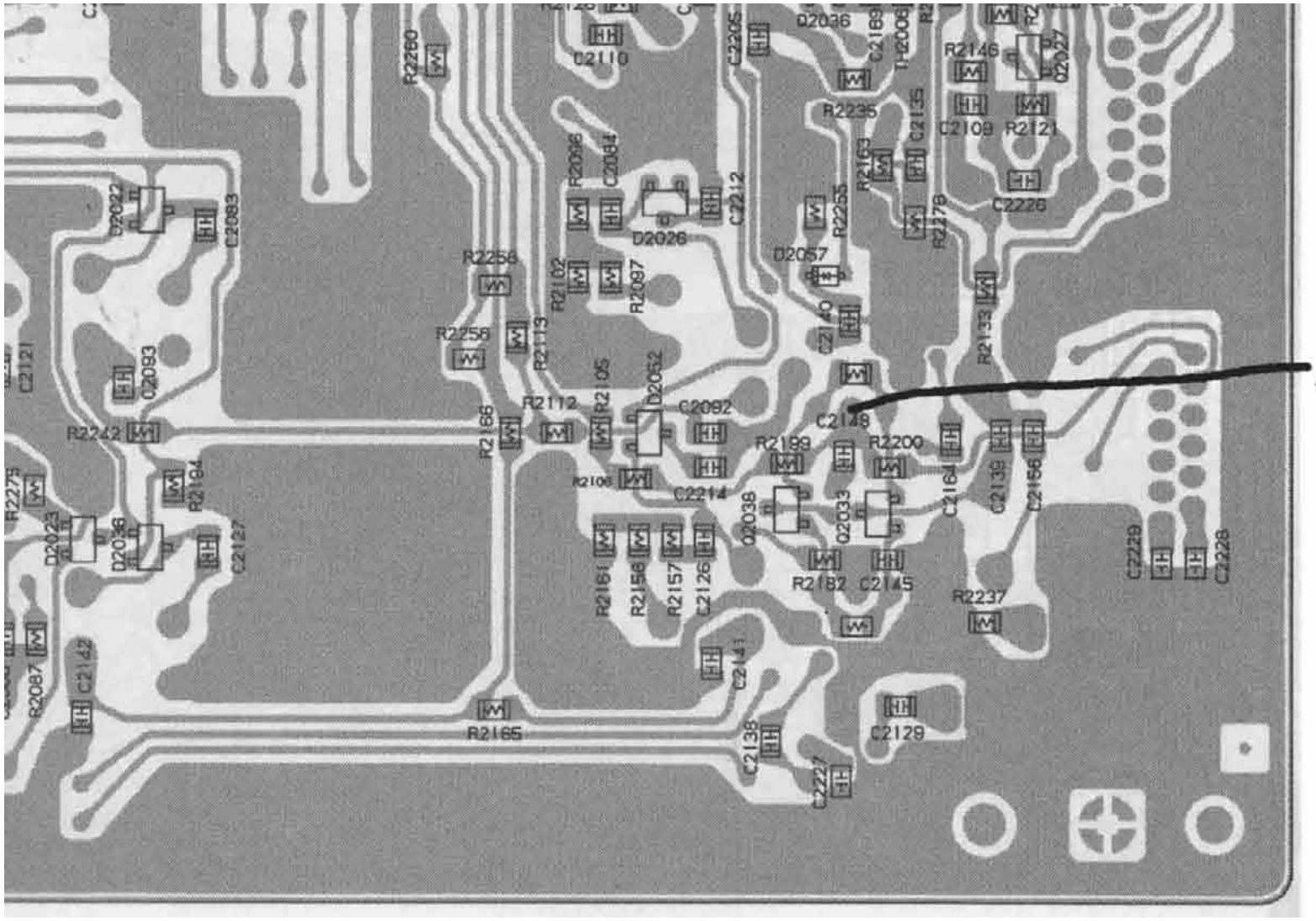
Reinstall the RF board and all other hardware taking care to not pinch any wires, and to reconnect all unplugged wires in proper locations

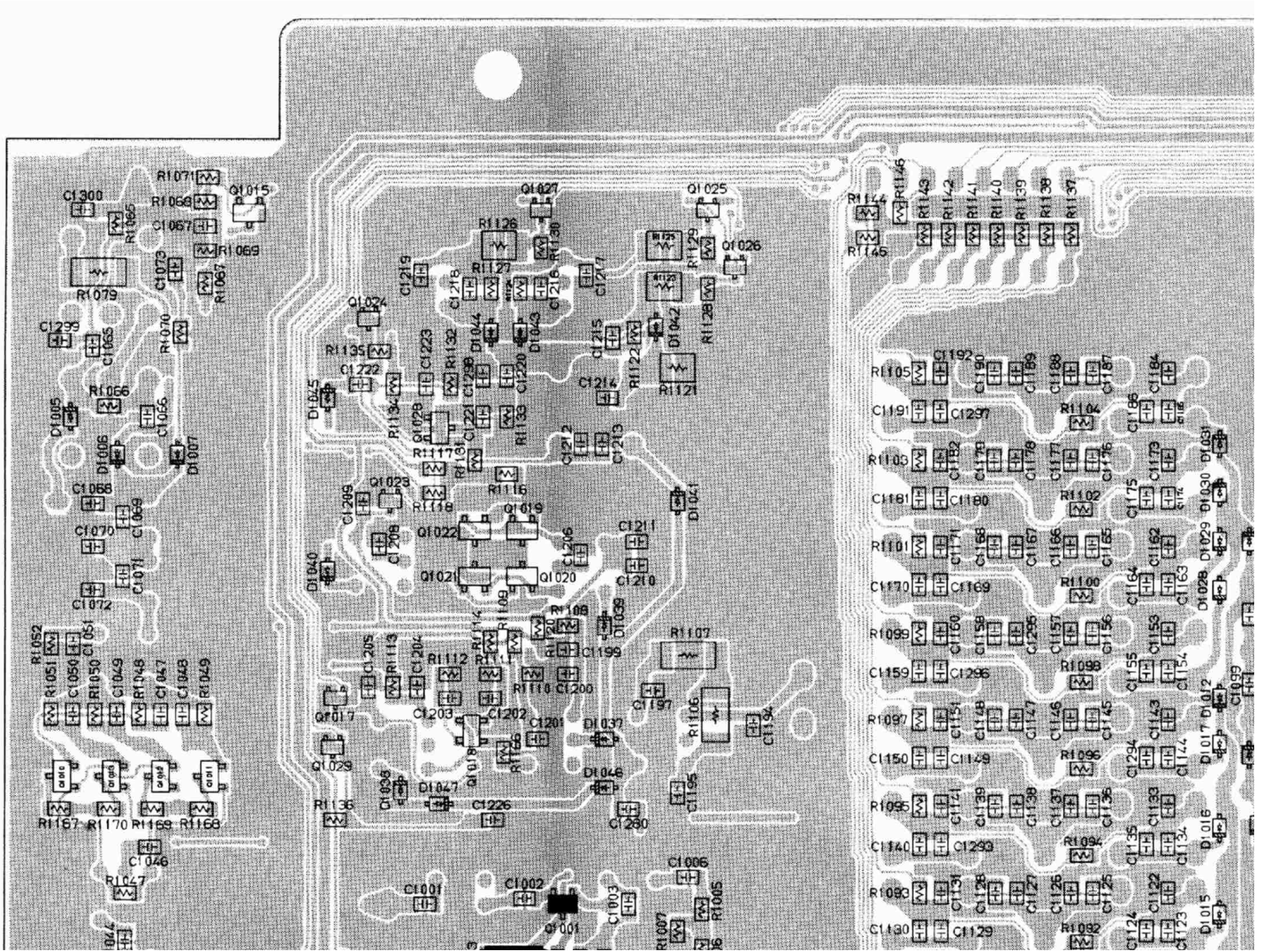
Select a clear area on the IF board and mount the terminal strip under the screws

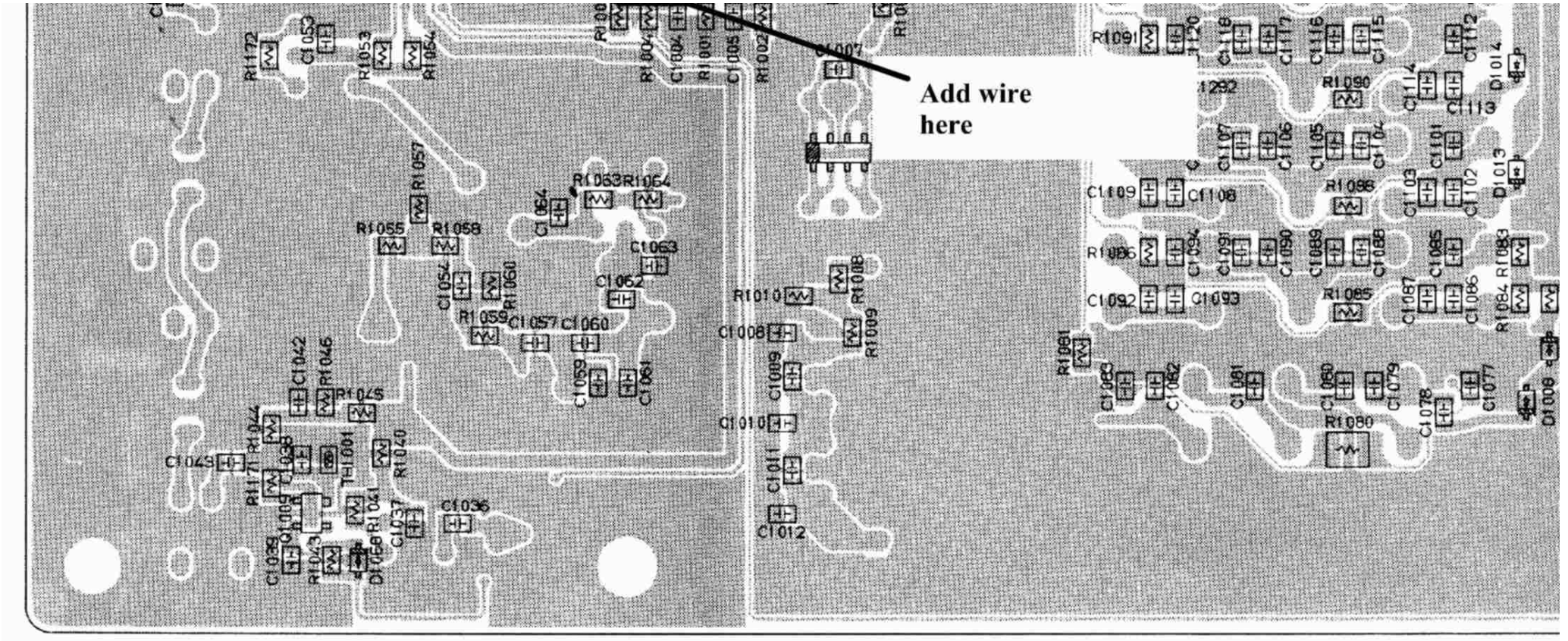
Connect the green wire (from RF board) to C3, and the blue wire from IF board to the junction of C1 and R1. R1 is a 680k resistor, and R2 is a 22k resistor

Test the radio and reinstall the covers


e
f
g
Remove R2046
add 220 ohm resistor
between these points


Add wire for click mod
C2148
C2110
C2109
R2121
C2226
R2146
R2235
C2135
R2163
R2255
R2179
D2057
C2140
R2133
D2026
C2212
C2084
R2086
R2087
R2102
R2113
R2256
R2258
C2121
D2093
C2083
D2022
R2242
R2166
R2112
R2105
D2052
C2092
R2199
R2200
C2164
C2139
C2156
C2214
D2038
D2033
R2182
C2145
R2237
C2229
C2228
R2161
R2158
R2157
C2126
C2127
D2023
D2036
R2184
R2275
C2142
R2087
R2165
C2141
C2138
C2227
C2129




Add wire
here

Keyclick_MP

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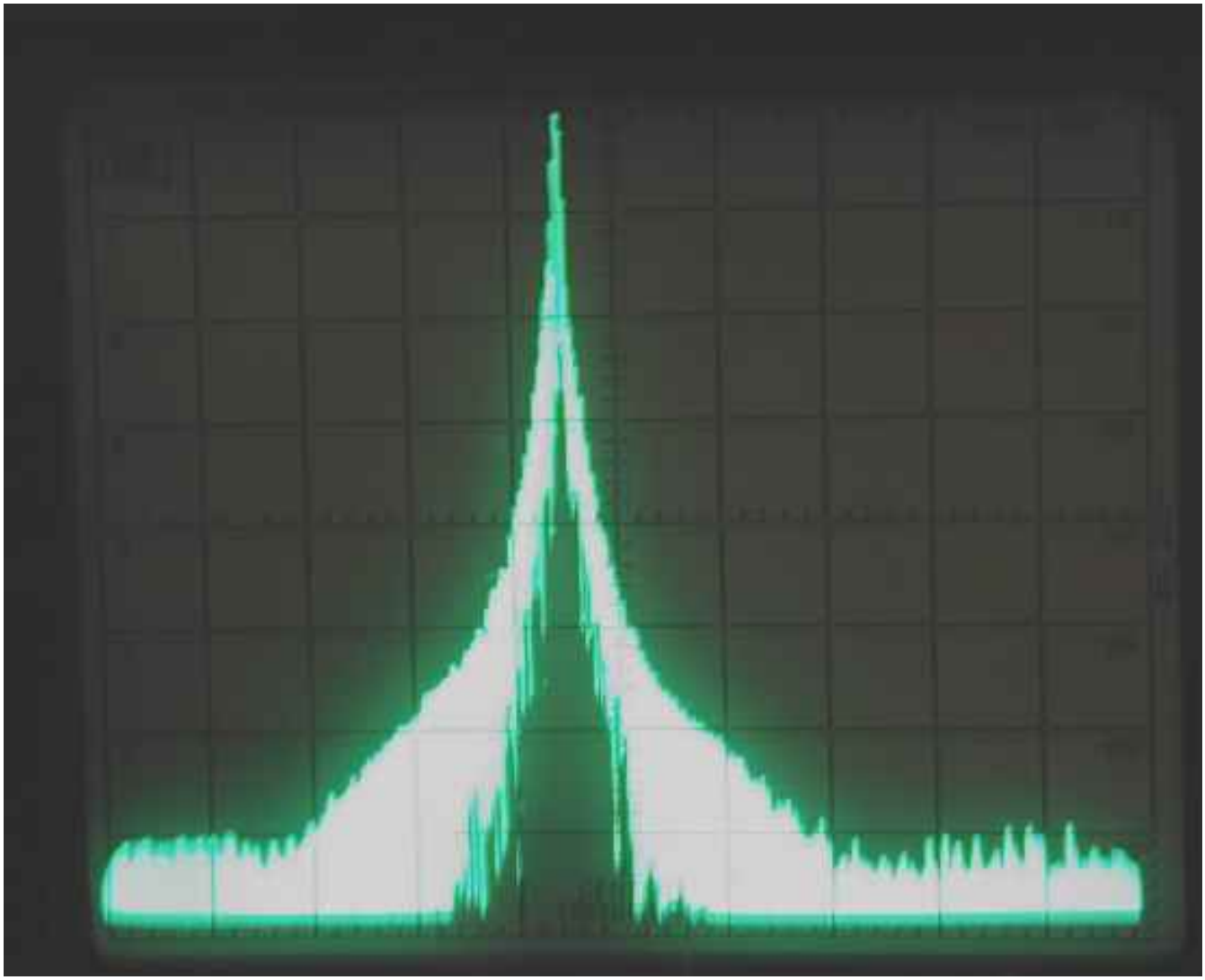
Curing FT 1000 MP CLICKS

FT-1000MP Measurements

I measured two sample FT-1000MP's (an early and a late model) by operating them into a high power fixed 30 dB attenuator pad. The output of the 30 dB pad was connected through a 3-way splitter to a step attenuator and conventional receiver, a spectrum analyzer, and an oscilloscope. The receiver used a 300 Hz eight-pole filter, the spectrum analyzer used a 50 Hz filter, and the scope was triggered from an external keying signal. Power was measured on a conventional Bird average reading meter.

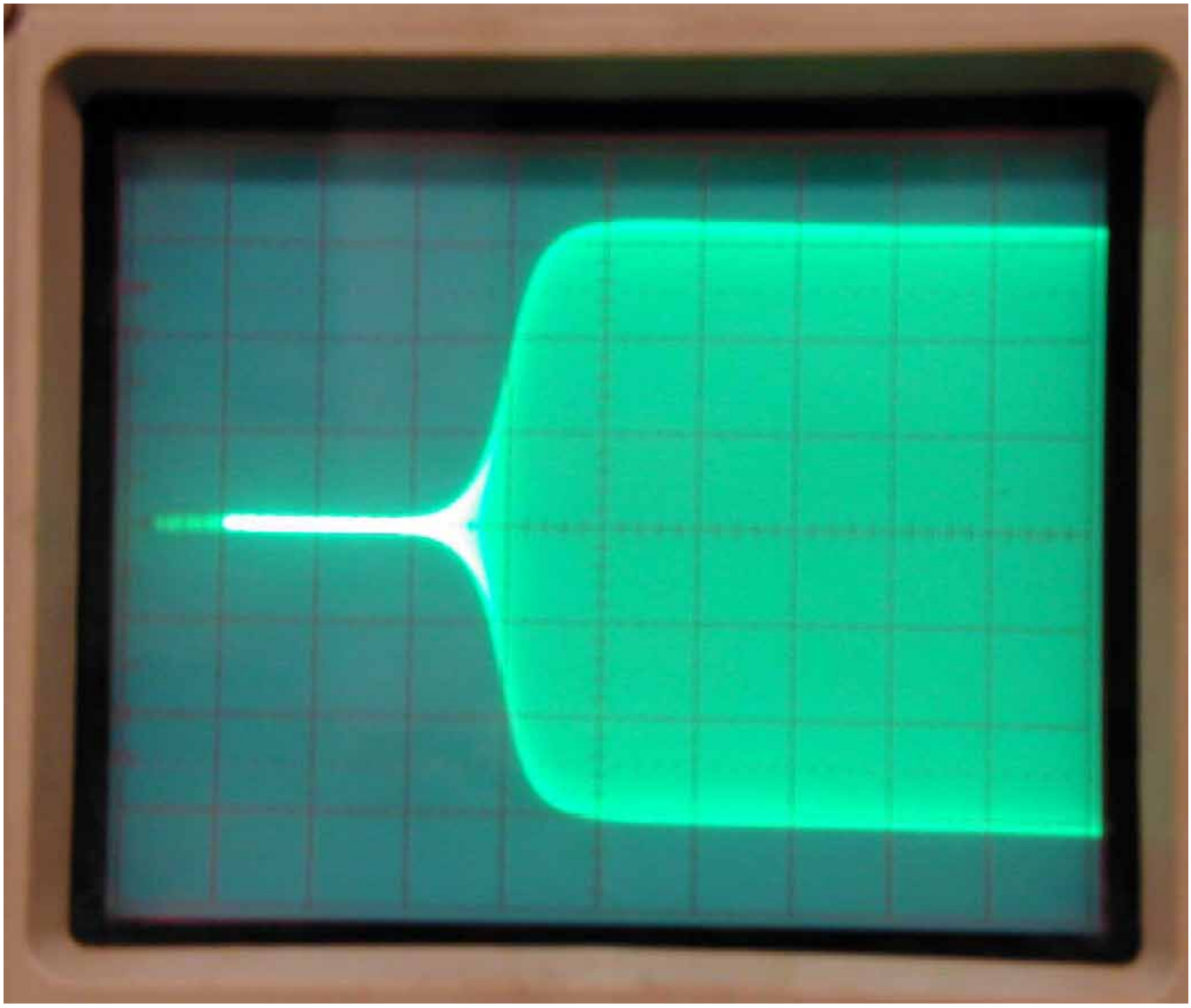
At 1kHz spacing clicks from the stock FT1000MP's were about 15 dB worse than clicks from my old test bench radio (a well-worn ICOM IC-751A) and more than 20dB worse than the clicks from my click-reduced FT-1000D!

Here is a spectrum display of my stock IC-751A using 30 Hz analyzer bandwidth and ten second sweep:

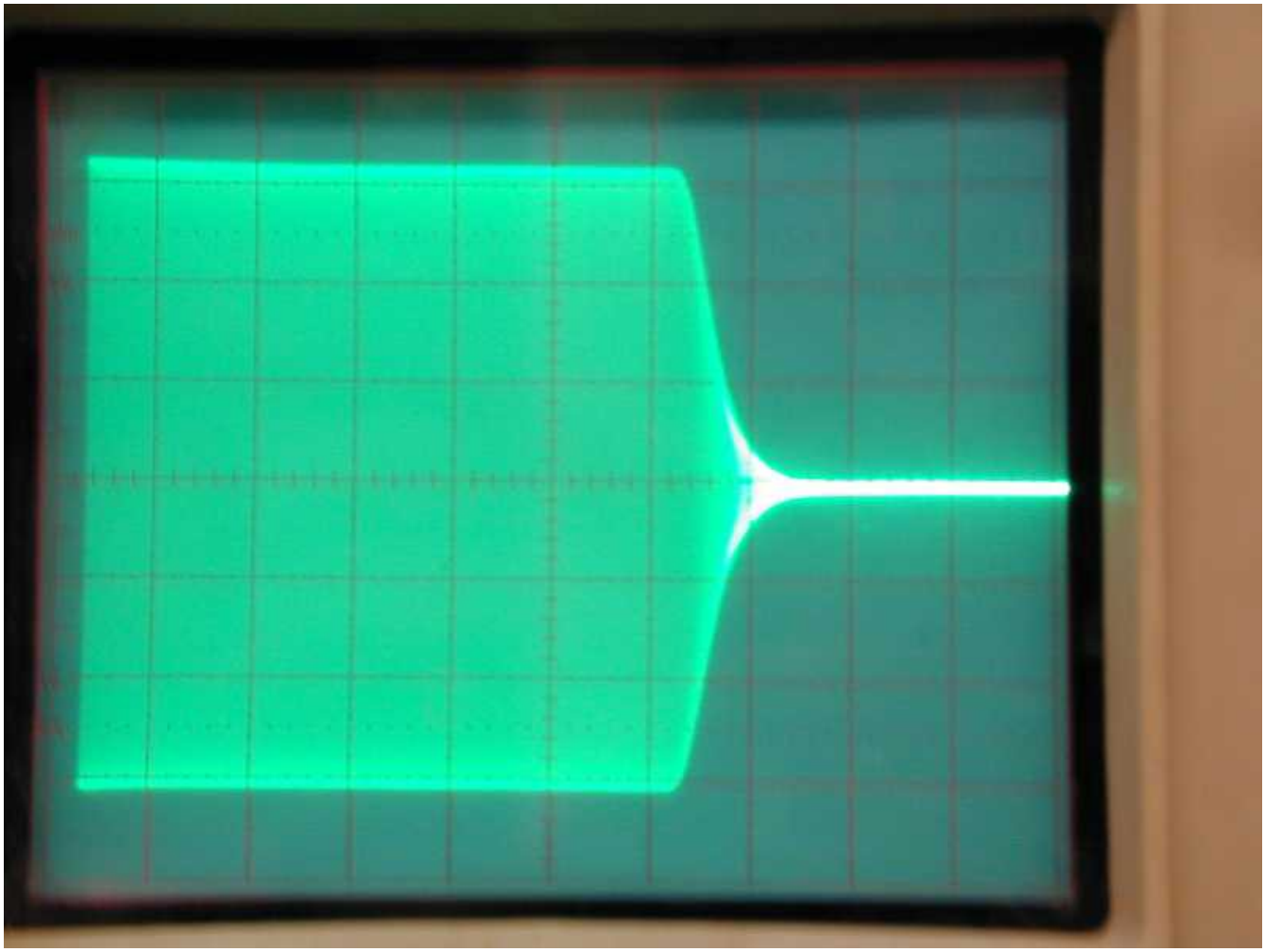


The 751A is approximately -58 dB at 1 kHz, and rolls off smoothly.

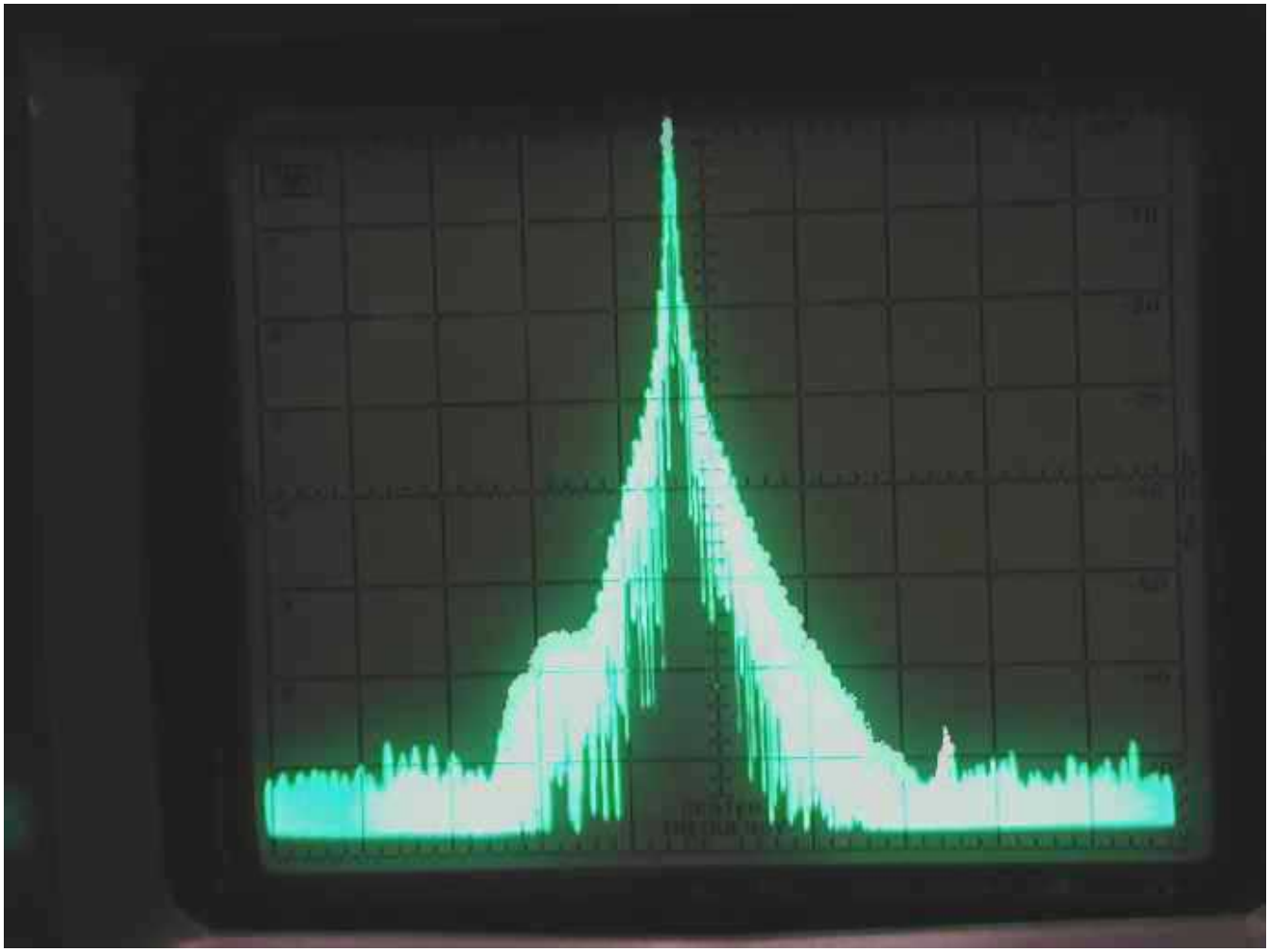
Here is the nicely sloped but too fast rise:



Rise approximately 3 mS....and the slightly sharper at the corner fall:

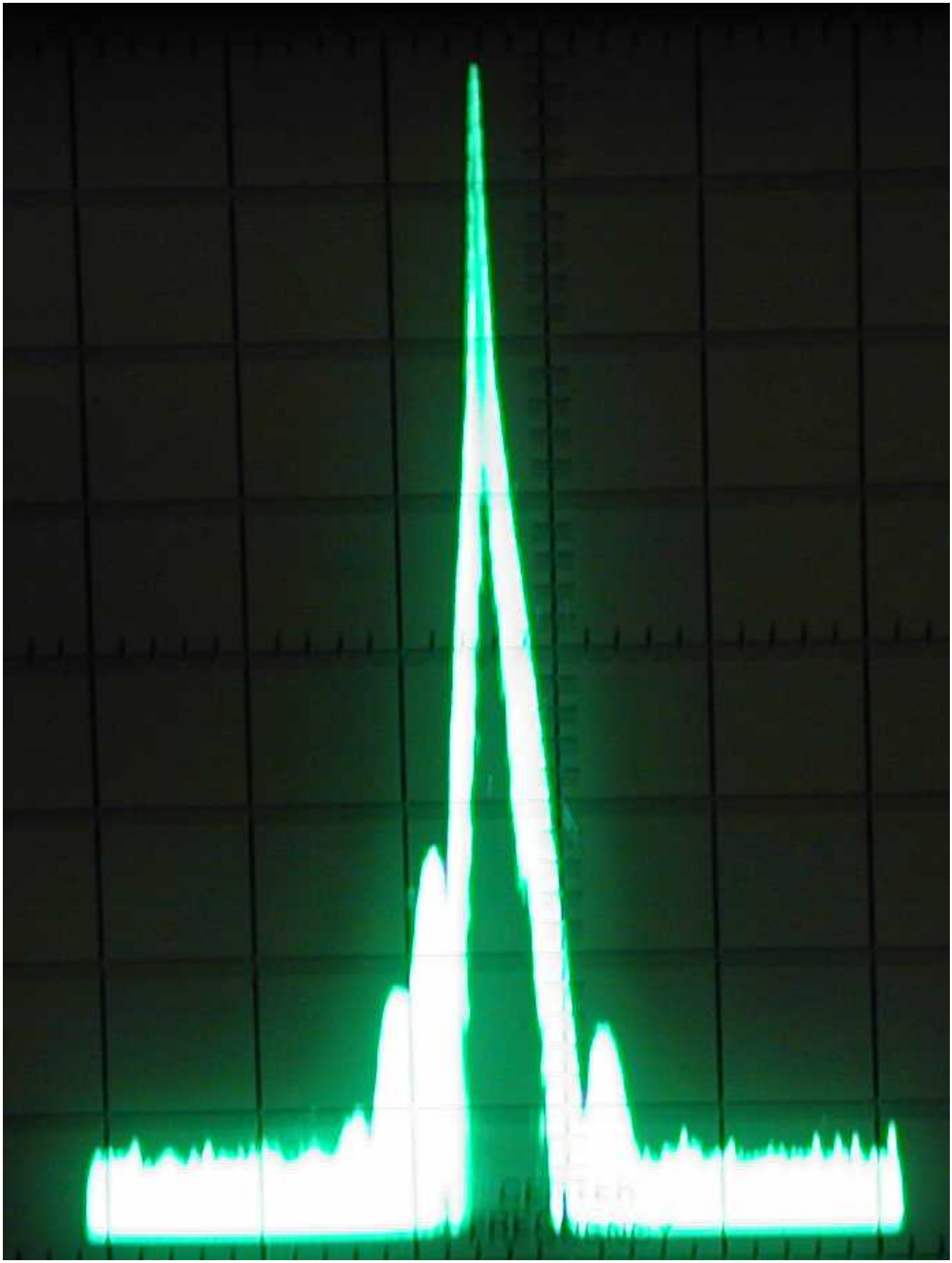


Now the stock 1000MP spectrum:



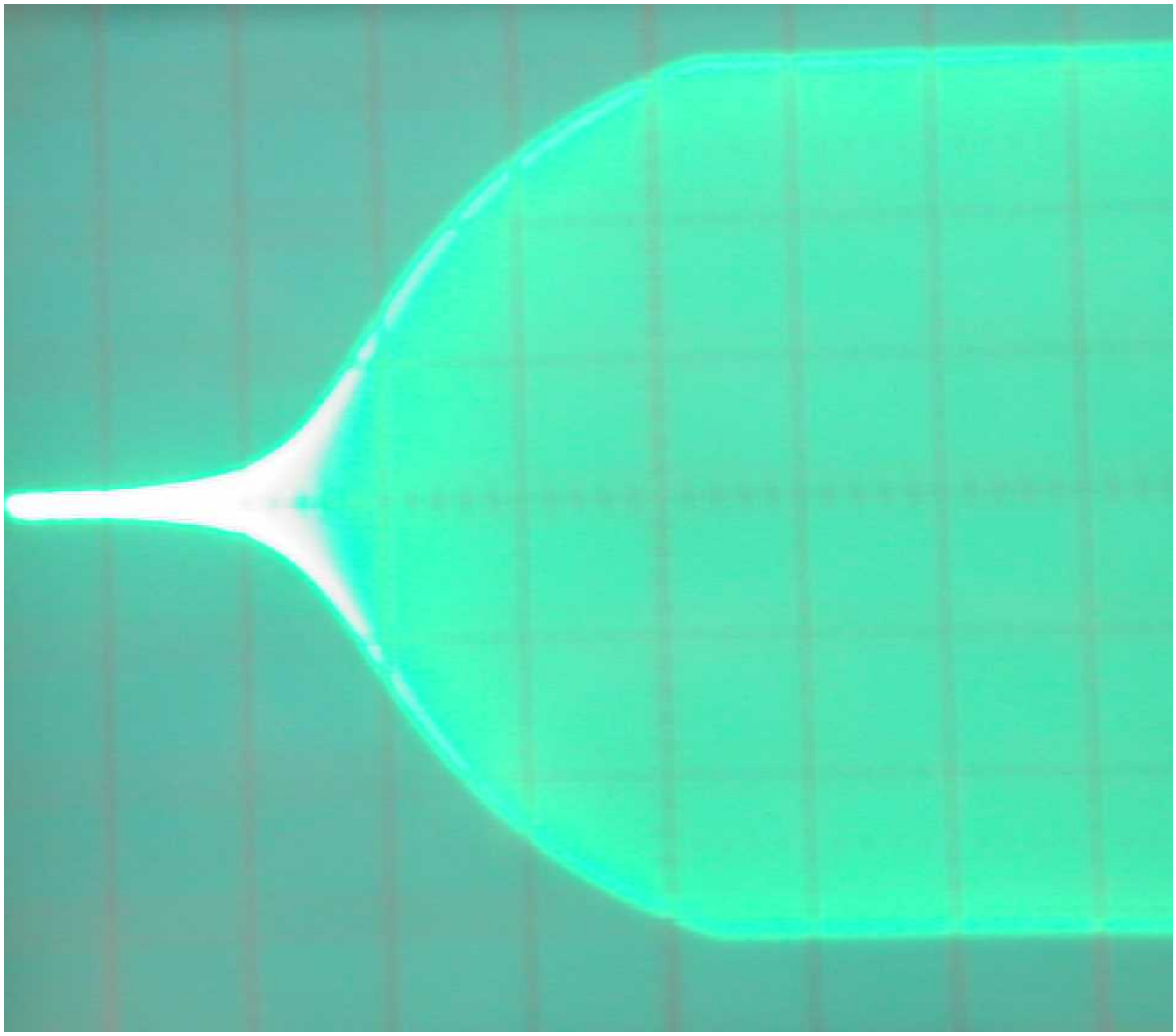
The FT1000-MP is approximately -50dB at 1kHz. It is 8 dB worse than the already "hard" 751A, and has a plateau below the carrier frequency that hovers around -55dB.

In direct comparison, here is a "de-clicked" FT-1000MP:



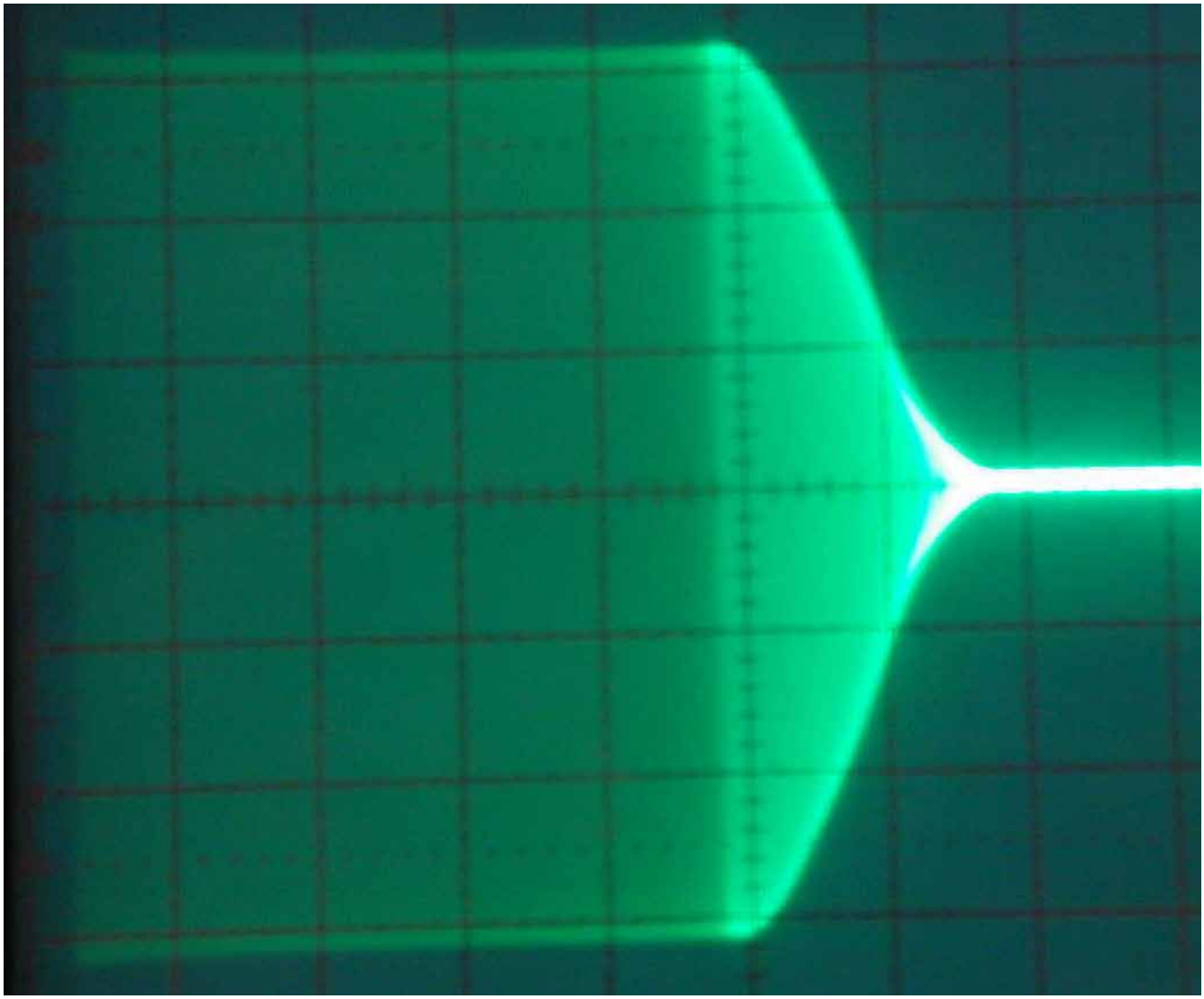
The modified FT-1000MP rivals any of the better radios I have tested, including my "de-clicked" FT-1000(D).

The modified FT-1000MP is around -85 dB at 1 kHz, over 30 dB improvement from the stock MP! Rise time is close to ARRL standards of 5mS, while fall time is around 3 mS. FT1000MP modified rise:



Rise 6 mS. The upper edge is a little sharp, but why worry....clicks are reduced 30dB or more!

Modified MP fall:



Fall time is around 3mS. While it has much more rounded edges, the slope is still not very "round". Unfortunately we are limited by what is possible to do, and this mod is already difficult enough for laymen.

Some concerned was expressed over the "power" of dots when using long rise and fall times. One simple solution is to turn up the weight control slightly. Keep in mind, even without ANY external weight adjustment, the change in average dot power at 45 WPM was only a few percent! On the air tests with VK3ZL and ZL3REX on 160 meters with fading signals and noise, revealed both could tell absolutely no difference between having the click filter in-line and out-of-line at 40 WPM CW speed. This waveform meets FCC and CCIR specifications for 60-WPM CW modest strength signals, and 35 WPM weak fading signals.

The best method of nulling clicks is by listening on another receiver with a narrow filter. Make sure you are well below overload on the receiver, and set that receiver so the carrier from the MP is just outside the passband of the test radio filter. It is almost impossible to use any other method for proper adjustment, including watching the envelope on an oscilloscope.

Work in a clear uncluttered location, I like to work with the radio on a clean small carpet on a well-lit bench, and have a container for all the hardware I remove.

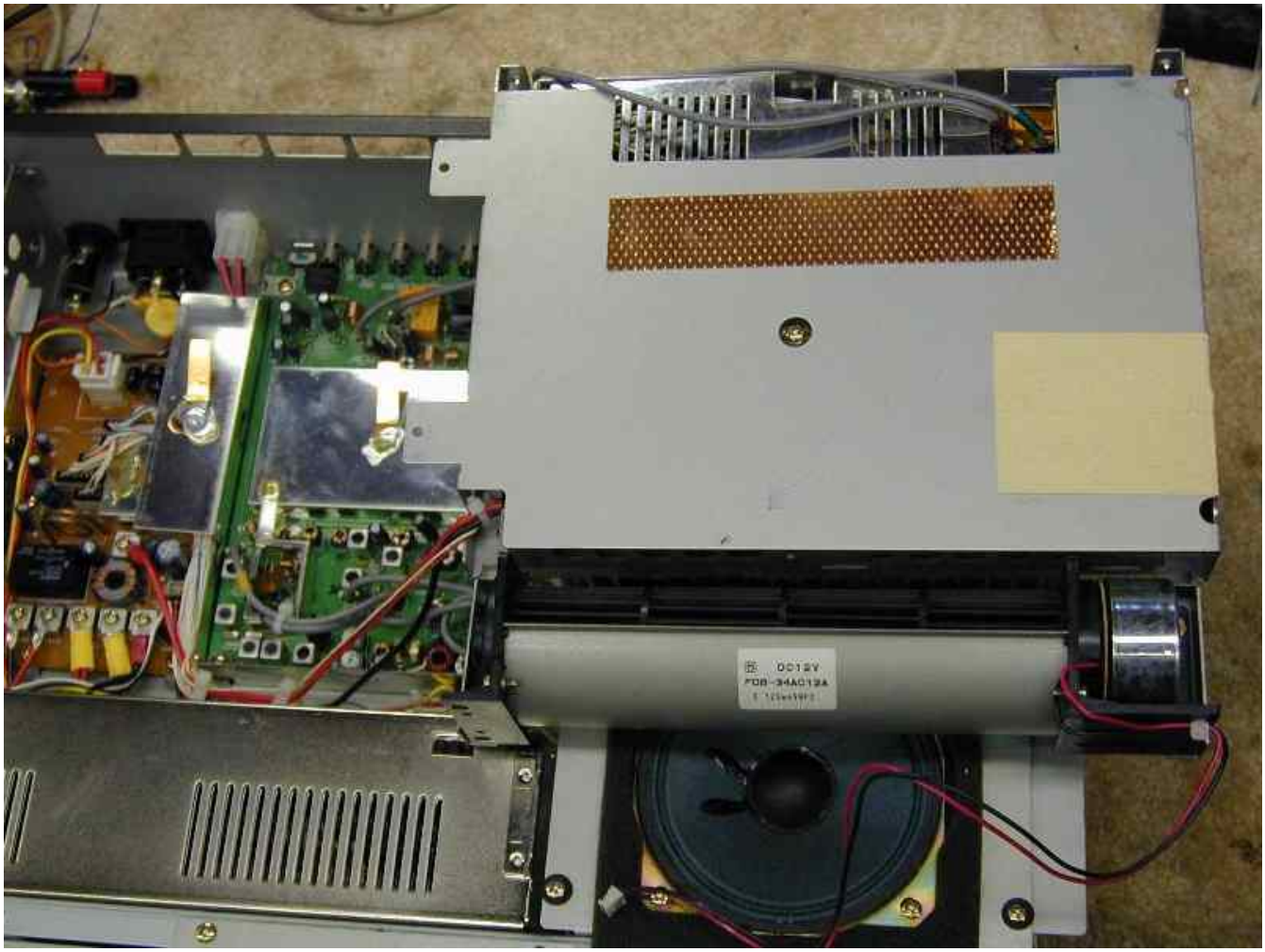
Here's how to make connections to necessary points:

1.) Remove the top and bottom covers of the radio and set them aside.

2.) Invert the radio, so you have the heatsink exposed.

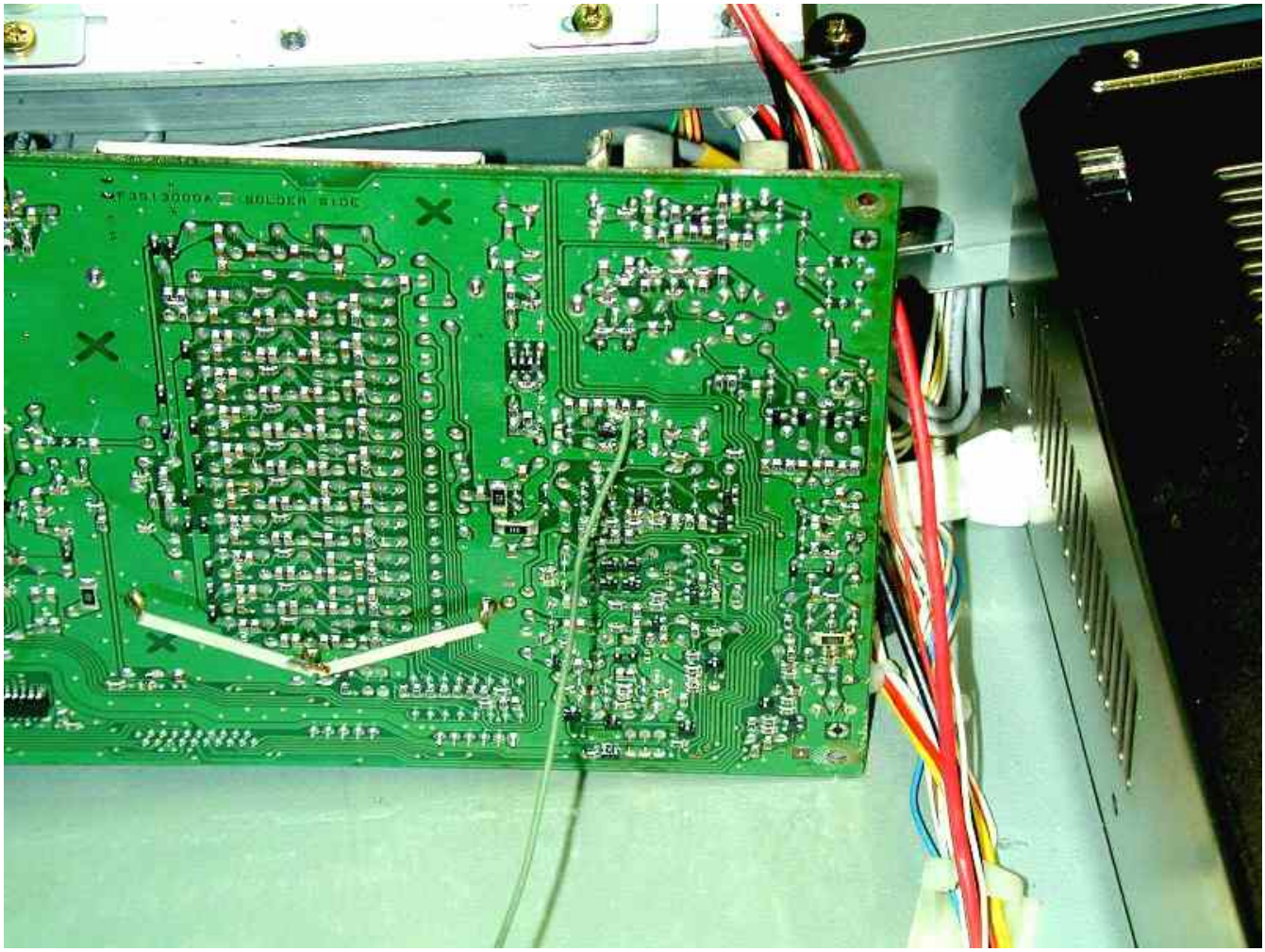
Four main screws hold the heatsink mounting bracket. Two are shown above (one under the screwdriver and one to the right of it). Two other screws are on the side of the radio chassis. You might want to remove the long screws holding the fan bracket, although I got by without doing so.

3.) Lay the fan and PA assembly out of the way, you may have to open some of the wire harness clips to get more wire. Unplug the fan so it is totally free from the unit. It should look like this now:



4.) The RF board is the green-colored board you see above. There are several screws holding that board down, and two screws on a rear panel DIN connector that is mounted on that board. The board will freely move when you remove ALL the screws. Do NOT pry or force the board out, if you have to pry you missed a screw!

5.) Flip the board over, you might have to unplug a wire harness or two...but I managed to work without doing that. The board should look like this:



6.) I added the green wire you see above. I tacked it on a foil pad by laying the wire across the point where two chip components soldered in, as shown below:

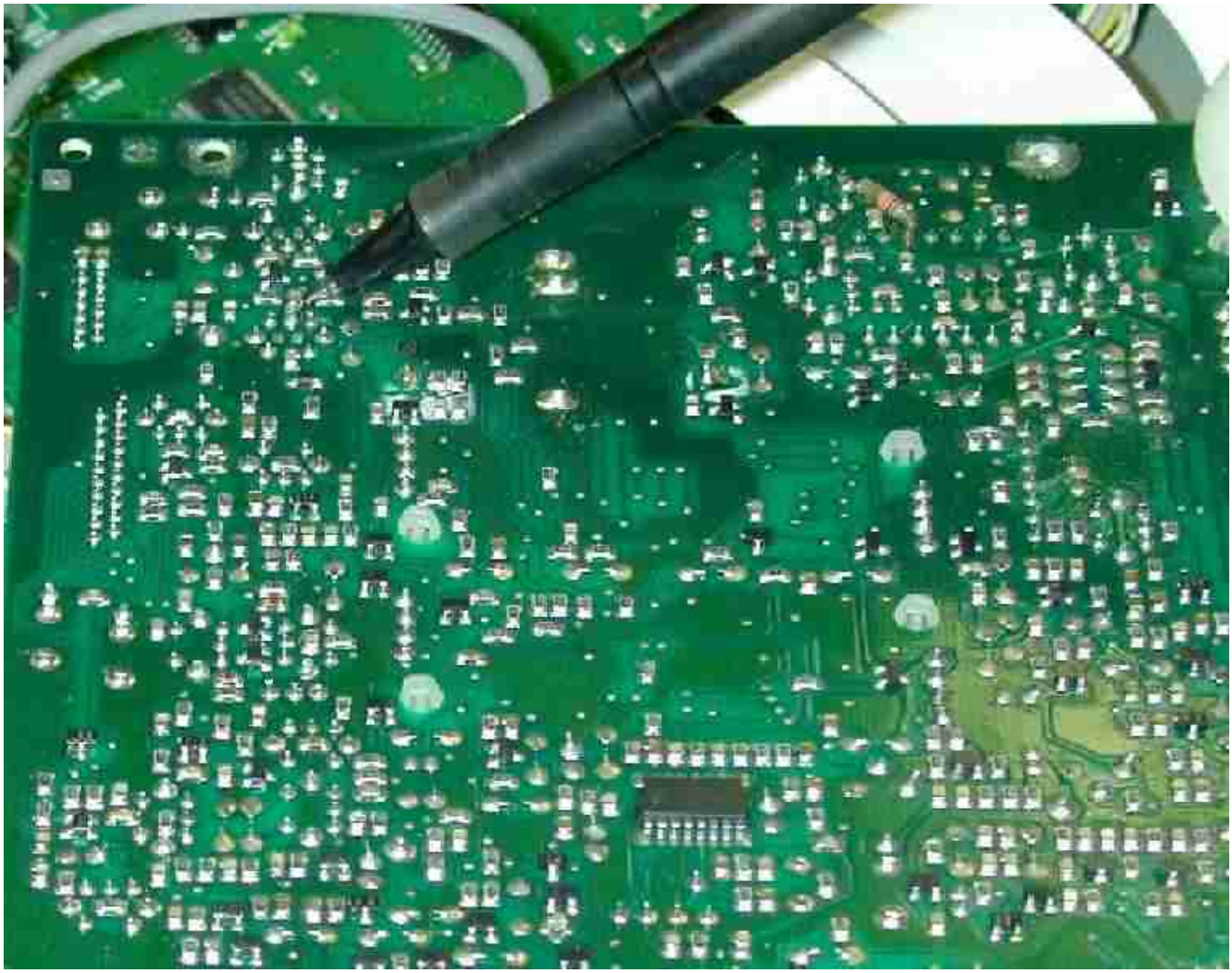
You can see the little black FET (Q1034) and the slightly triangle-shaped foil trace that connects to Q1034. I bent an "L" in the small wire I used. The wire will route topside to a terminal strip, so it needs to be several inches long.

7.) The new wire routes under the RF board to an oval slot in the chassis, where if feeds to the other side and emerges near the IF section unit. Be careful not to pinch any wires when remounting the circuit board. The opening on the right is best for getting the added wire topside:

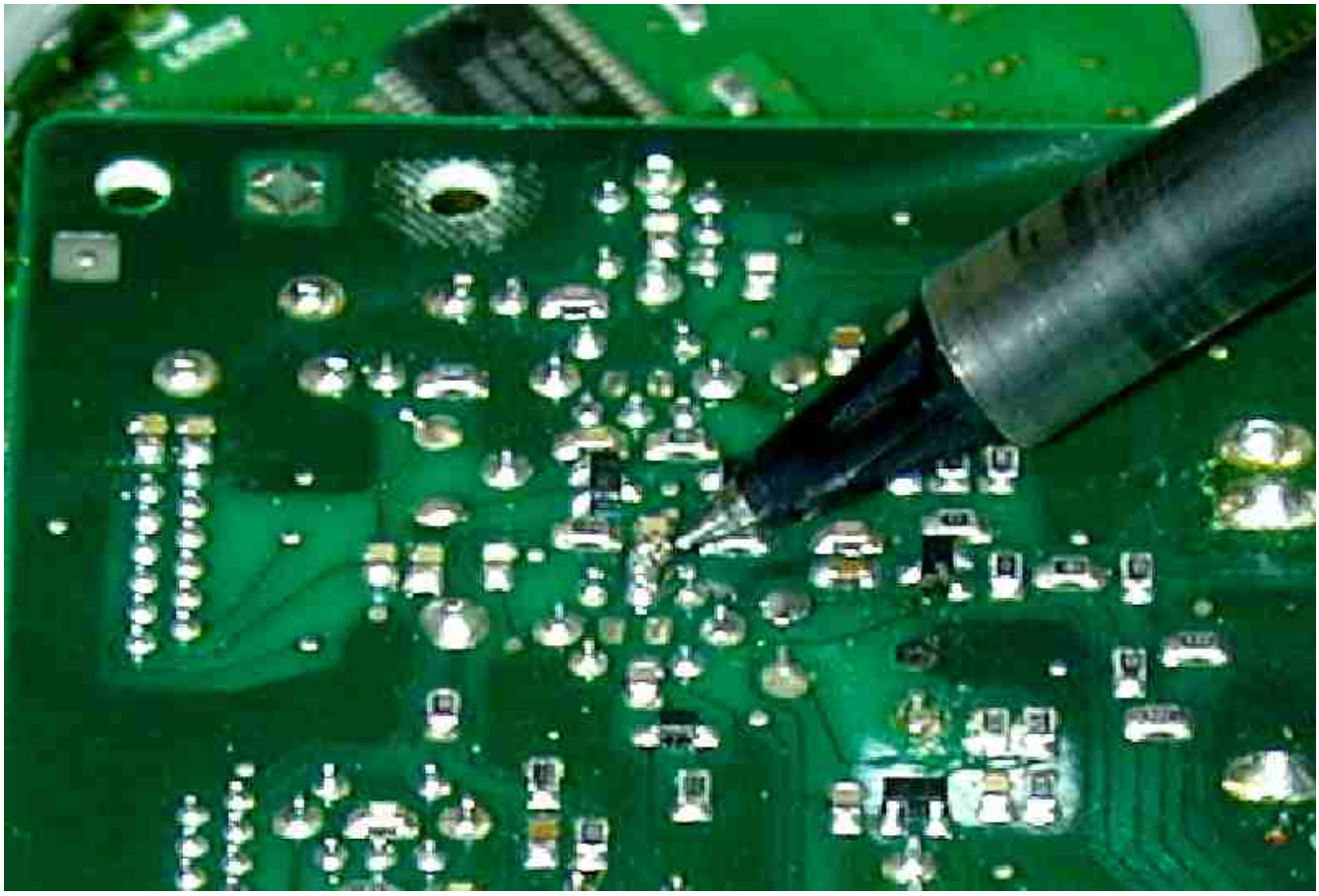
8.) While re-installing the RF section, fan, and heatsink inspect the wiring carefully. Be sure nothing is touching moving parts of the fan, and be sure no wires are pinched or left unplugged. It might be advisable to check the radio quickly on a dummy load to be sure the transmitter section works properly.

9.) Flip the radio over, and remove the two ribbon cables connecting to the IF section:

10.) Remove the mounting screws and flip the IF board exposing the bottom:



This is the area where the wire attaches to the IF board:



Note the FET above and to the left of the pen, and the IF transformer (two can shield connections and five electrical pin-outs, with an unused pad-set for a surface mount device in the center of the transformer leads) below the pen. The connection point for a new small insulated wire is the chip capacitor (C2148) pad that also connects to the center pin of the top three in-line pins of the IF transformer.

11.) Attach a small wire to the point mentioned above. It will route to the new circuitry.

12.) Re-install the board. Be sure you do not pinch any wires. Be sure you do not forget to plug in the ribbon cables, or any other wires you removed.

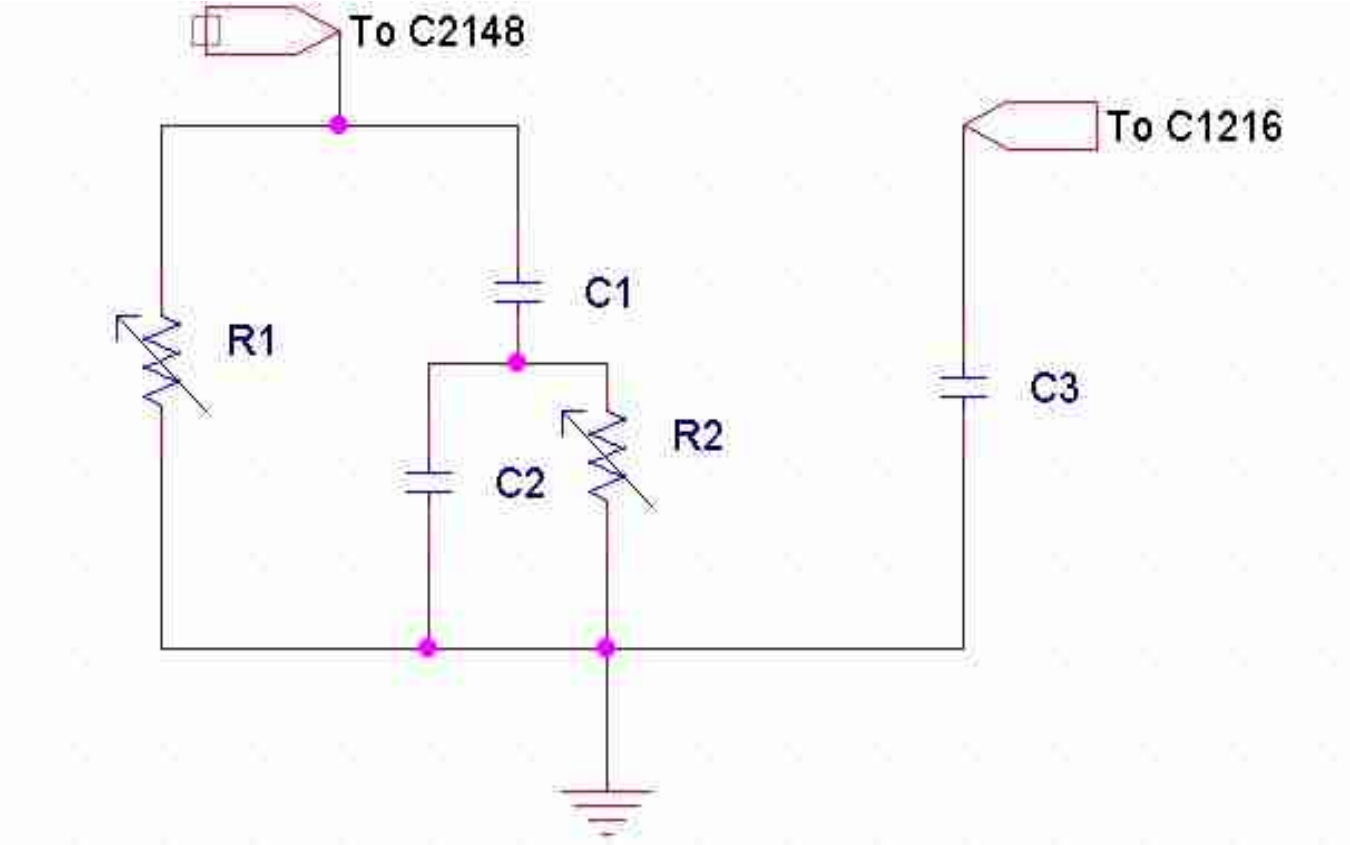
I mounted a small four lug (plus ground) terminal strip at the left front corner of this board, and mounted the components on that strip. This allows you the ability to change or adjust the mod if you find it too soft for high-speed operation.

Doing The Mod

The ideal CW radio would use a high-order filter with controlled group delay, and a reasonably linear attenuator or modulator to control the envelope shape. All other stages should be fully on just outside the output window of the CW signal. I initially hoped a CW "modulator" could be

added on, but for now it appears modifying the 1000MP to ideal circuitry would be too involved. My only option was to "hunt and peck" and find a modification that would be reasonable to do, and inexpensive. This is the best solution I could find, reduction of clicks was excellent. The only drawback is two resistors need to be hand-selected, and the radio needs some minor disassembly to reach a connection point on the RF board.

I mounted this mod on a separate terminal strip under a screw on the left front corner of the IF board. This allowed me to experiment with component values while watching bandwidth and other parameters. This is the basic circuit I used:

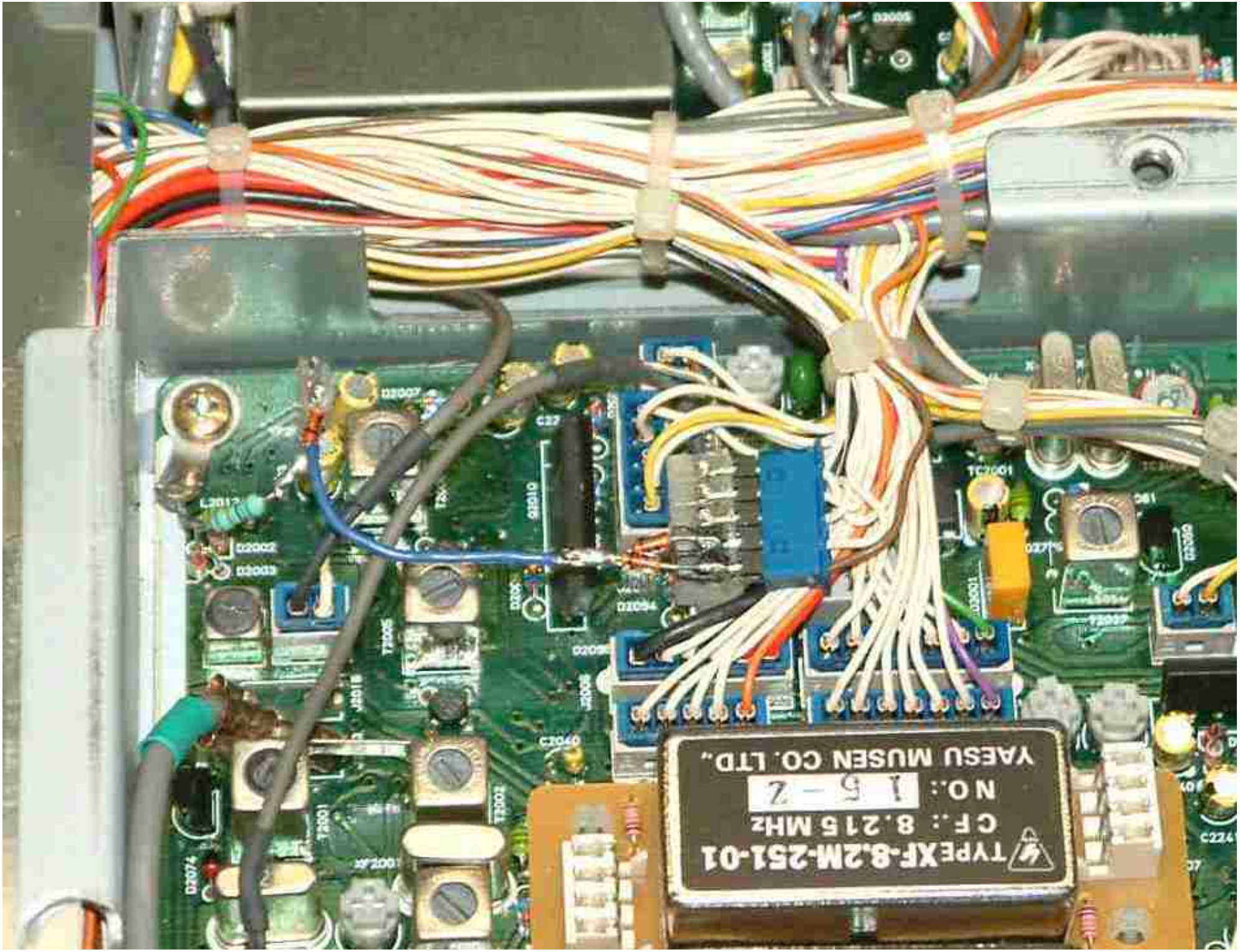


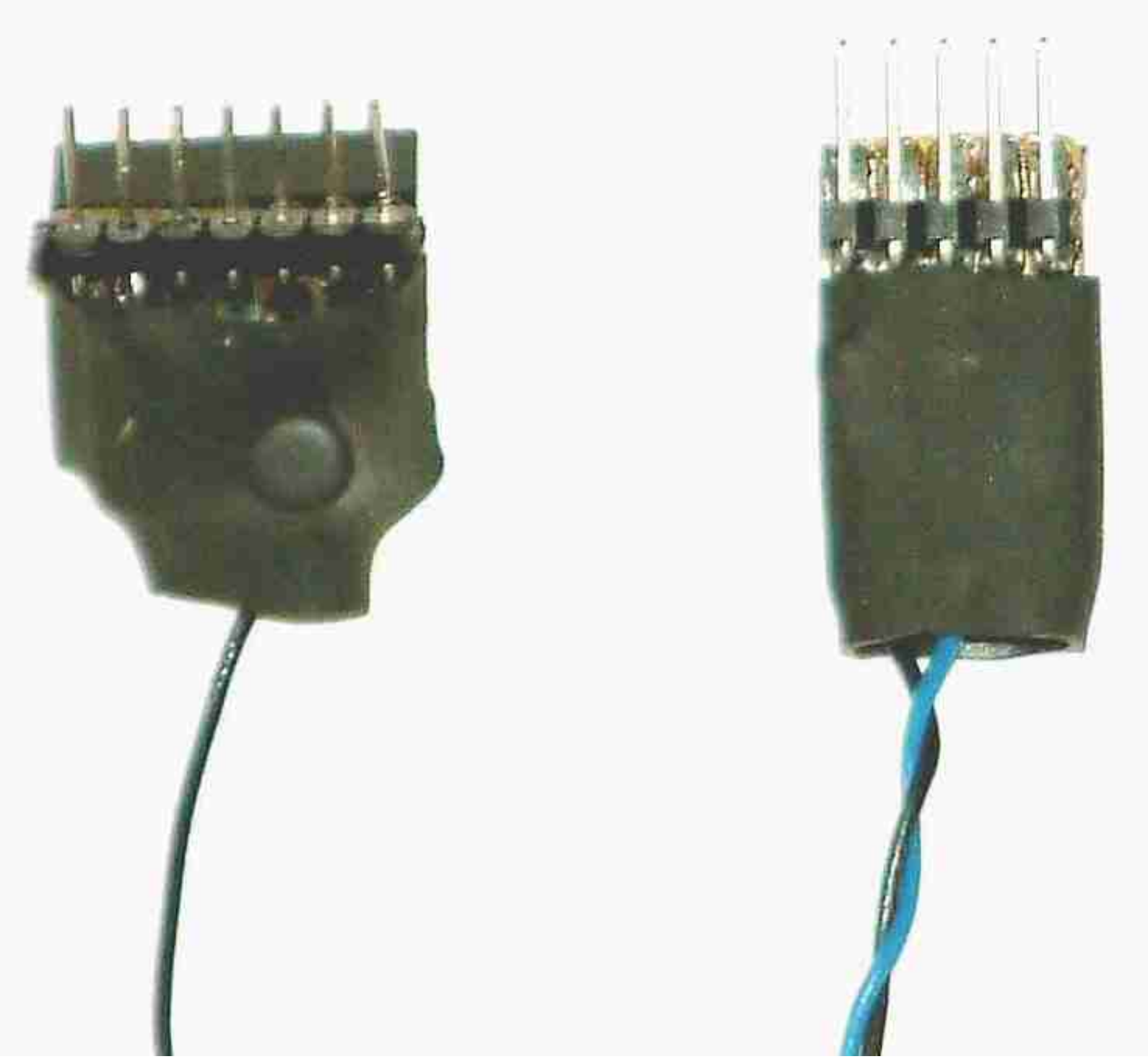
C3 was a .1uF disk capacitor. The value turned out to not be especially critical, it mainly seems to prevent rapid rise and fall of the low-level RF amplifier stage that is driven by a gate. There was no combination of resistance across or in series with this capacitor that reduced clicks in any of the radios I tested.

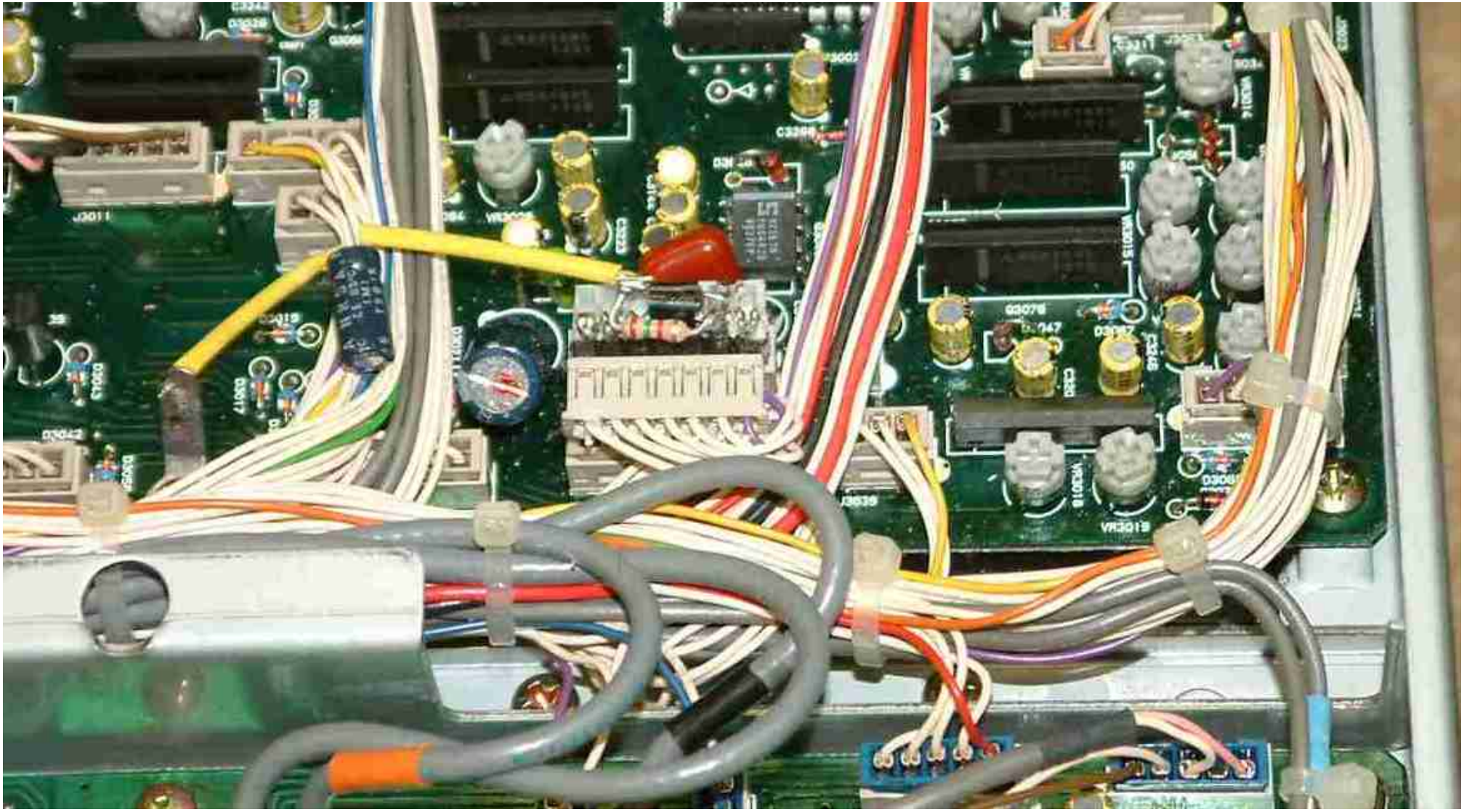
C1 and C2 are also .1uF disc capacitors. In all units tested, I could find no better combination for reducing clicks.

The only critical components appear to be R1 and R2. R1 ranged from 120k to 470k in the units I tested. R2 ranged from 1k to 10k ohms. I initially clipped in potentiometers, so I could listen to the

output and adjust the clicks at 1kHz spacing. Both pots were adjusted for a null in click amplitude. That null is rather sharp, and turned out to be around 30 dB deep. This takes the 1000MP from being one of the "clickiest" radios I have found to one of the cleanest!


TYPE XF-8.2M-251-01
CF: 8.215 MHz
NO.: 1 5 - 2
YAESU MUSEN CO. LTD.





What_Causes_Clicks?

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Bandwidth rules Part 97.307

Clicks are often a problem on congested bands, yet with modern radios they don't need to be a problem. In the past, engineers and designers didn't have the easy ability to generate filtered waveforms. Radios lacked wide 2-3 kHz wide SSB filters, let alone narrow 250-500Hz filters. Op-amps were unheard of, and L-C filters were large, bulky and expensive. Today, every radio manufactured has the ability to be very clean.

Modern Radios

Most modern radios include 500Hz and narrower receiver filters that operate at the same IF frequency as their transmitter section. Transmitted signals are often routed through the SSB filters with intentional TX switching, when they could just as easily be routed through CW filters! We will see later in this page that TX signals could be routed through CW filters to eliminate sidebands, yet manufacturers foolishly use the SSB filters. If you examine the bandwidth of a FT1000-series radio, you will find the clicks disappear at the BW limits of the SSB filter. This is because they run an essentially unshaped CW signal through the SSB filter, and that filter sets the bandwidth of clicks.

The sad thing is once you listen to that signal through a 500Hz filter, it sounds absolutely no different ***ON FREQUENCY*** than it would if it were nearly click free. The only people who can notice the difference between a clicking rig and a clean rig are the people operating on adjacent frequencies! Claims that certain shapes produce certain "bell sounds" or high readability are not correct, they are certainly not based on engineering or actual blind A-B tests.

If you examine the audio output of your radio with a 500Hz filter in use, you will see ANY waveshape transmitter has the same nearly raised-sine shape output to your ear! That includes wide signals and narrow ones. The speed limit you can copy with a 500Hz filter is the same limit you can transmit with. It makes no difference what end of the path the filter is on, or if there is one at both ends, so far as speed is concerned! (This assumes the filters have reasonably good and easy to achieve group-delay characteristics.) We hear a few mS rise, no matter if it is a sine shape or a square, as a "tick". For demonstration, listen to the pure sine wave on WWV that "tics" every second!

What Could Be Done

At no cost to manufacturers, they could build a click free radio. Every component is in the radio, the problem all centers around poor or careless engineering.

Amplifier stages are reasonably linear (so they can amplify SSB), and virtually every radio contains power control circuitry that could be easily modified to provide wave-shaping. Even without wave-shaping, the transmitter could process transmitted CW though a 250Hz or 500Hz filter.

Sadly, most of the commonly used radios have as bad or worse keying characteristics than old rigs. It's as if the manufactures either don't understand CW, or don't care. The result is we are left with a mess, because many top-of-the-line and very popular rigs have horrible keying sidebands.

On frequency with normal CW filters, we would not be able to tell any difference between the sound of a clicking radio and one that is clean! There is no justification or reason for radios to be 3kHz wide on CW.

How to Identify Click Problems

We hardly notice clicks, and we certainly can not tell a clean rig from a dirty rig, when we are listening right on the CW station's frequency! Even an scope won't tell us much about signal bandwidth, or if the rig has excessive clicks.

In order to check clicks, we must:

1. Be sure the receiver is not overloading
2. Listen with the CW signal outside the receiver filter's bandwidth
3. Listen when the noise is low, and the signal reasonably strong

If we do not follow those three guidelines, we can't tell if a rig is clean or not. If you are testing your own rig, your second receiver must have a narrow filter and be coupled to the rig-under-test through a proper attenuator.

Why Worry About Clicks?

Clicks are most problematic when we try to copy weak signals next to moderately strong signals. If you only operate on empty bands, run low power, and never operate within four or five kHz of weak stations, bandwidth is probably not a concern.

If we contest, work DX, or Ragchew near other QSO's, and especially when we run more than a few hundred watts and have large antennas, we should be mindful of our bandwidth. If you listen to a

recording of a clicking radio, you can hear how devastating clicks are to nearby weaker signals. This signal is from Europe on 40 meters, and it is daylight over half of the path!!

For a mathematical tutorial on clicks, visit W9CF's site. Kevin's analysis deals with bandwidth requirements related ONLY to modulation of the envelope. I'll explain the same thing in verbal form, as I discuss sidebands created by rise and fall times. CW keying is really just 100% AM modulation, as you will see!

There are several INCORRECT but popular misconceptions. They are:

- **A signal has to be clicking and/or wide to send fast CW**
- **Clicks or sharp rises and falls aid in weak signal work**
- **Your CW signal bandwidth changes with the speed you are sending**
- **A certain shape gives a certain on-frequency sound**

What Causes Clicks?

While a fast rise and fall time guarantee excessive bandwidth, ***a long rise and fall is no guarantee a radio will be "click-free".*** Some radios switch into transmit while the synthesizer (VCO) circuits are still settling to a new frequency. An IC-775DSP I owned was particularly bad about this, and also had VCO leakage problems. The amount of garbage varied with how I used the radio, including "VFO" frequency settings of unused VFO's!

Radios with VCO or synthesizer settling time problems generally produce a loud "thump" on key closure on the second VCO frequency. That thump will be right on the DX station when the operator is working split. If you listen in pileups, you will hear a small percentage of rigs with this problem. If the operator uses QSK, VCO-switching-thumps can be particularly annoying. Thumps will occur every time the VCO moves from the receive frequency to the transmit frequency, sounding like a leading-edge click!

Rise and fall times are also important. ***A long rise and fall time does not always result in narrow CW transmitter bandwidth, even though a faster-than-needed rise and fall time almost certainly results in excessive bandwidth.*** Many radios have rise and fall times that are much too fast.

How fast is much too fast? For now let's ignore VCO switching problems, and consider envelope shape.

Rise and Fall

The ARRL recommends a 5 mS rise and 5 mS fall time for CW, based on data in section 2.202 of FCC rules and CCIR Radio regulations. According to professional sources, a 5 ms rise and fall time is not harmful to readability at 35 wpm under marginal (fading) conditions, and 60 wpm when signals are reasonably above noise floor. This rise and fall results in a occupied bandwidth of 150 Hz, although unwanted transient energy caused by the shape of the waveform slope may appear at wider bandwidths.

What Limits Bandwidth?

When determining bandwidth of a stable signal (no oscillator problems), two things come into play in.

- The ***slope*** of the envelope rise and fall at any point controls the band***width*** of the keying sidebands
- The ***amount of voltage change*** during the slope controls the power ***level*** of the sidebands (clicks).

The slope (bandwidth) and the amount of change in a sloped area (level) combine to determine how offensive the transmitted signal is. ***Very subtle changes in envelope shape have a profound effect on key click amplitude and frequency dispersion.*** This makes it nearly impossible to tell if our radios are as clean as they could be by looking at envelope shape.

We can be certain sharp transitions will cause problems, especially if we can actually see them on a oscilloscope. We can also be sure that a rise and fall faster than 2 or 3 milliseconds will cause a bandwidth problem.

Reference Data for Radio Engineers, in the section of Radio Noise and Interference, addresses key clicks in a manner the ARRL Handbook does not. They give an example of multi-pole shaping of waveform. The ARRL Handbook seems stuck with the incorrect notion that a single-pole R/C filter provides proper shaping, something doubtless left over from 1940's technology when better filters were expensive, large, and complicated.

Here are the bandwidth curves of three basic envelope shapes, one rectangular (some radios are this bad!), one for a proper single pole R/C filter with slightly rounded shape (The ARRL suggests this shape. Probably because it was practical in the early years and "stuck" even though it is not ideal), and one for a filtered rise and fall (this would be a sine-shaped rise and fall from a multi-pole filter). We can clearly see a large difference in bandwidth in the curves below:

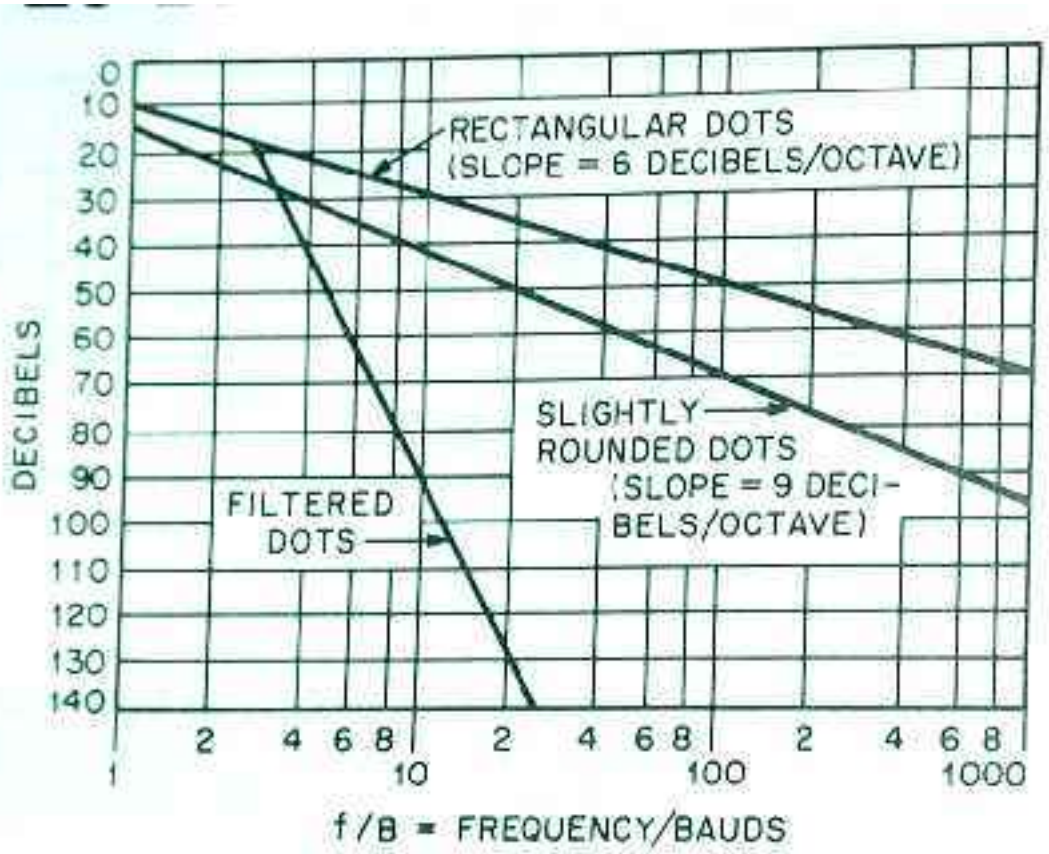

Fig. 12—Curves giving the envelopes for Fourier spectra of the emission resulting from several shapes of a single telegraph dot. For the upper curve the dot is taken to be rectangular and its length is $\frac{1}{2}$ of the period T corresponding to the fundamental dotting frequency. The dotting speed in bauds is $B=1/t=2/T$. The bottom curve would result from the insertion of a filter with a pass band equal to 5 units on the f/B scale, and having a slope of 30 decibels/octave outside of the pass band.

From Ref Data for Radio Engineers 29-10 1977 Edition

Most radios, through poor design, fit in the rectangular to slightly-rounded category!

What Can Manufacturers Do?

Radio manufacturers can certainly do a great deal more than they are. First, they created the problems through poor engineering and design. Why are we stuck fixing them? Did they take our money and run?

All of the parts are there to make radios virtually click-free, yet the only manufacturer who has taken an active interest in this (and who seems to care at all about our signal quality and frequency usage) is Ten-Tec! To date I haven't found any other manufacturer admitting a problem, or even offering technical support for bandwidth problems.

Let me give an example of what could be done with current radios:

Virtually every radio contains a CW filter that operates at the IF frequency of the transmitter, yet nearly every radio transmits CW through the SSB filter! Engineers actually ***added circuitry and parts***, in many cases, ***to steer the CW through the wider filter on transmit!*** If you listen to radios, in particular the FT1000-series, you will notice they have an ultimate click-bandwidth of about the same width as the SSB filter. That's because the poorly-shaped CW waveform with excessively fast rise-and-fall is filtered through the SSB filter.

If these same radios immediately turned on the output stages, and held them on for several mS after the key line was opened, they could send perfect filtered CW through the CW filter. A 500Hz filter would cause a steep roll-off in clicks, even if driven by a relatively "square" and very broad CW signal. The resulting waveform would be a slightly modified raised-sine envelope.

The listener would not be able to tell *any difference* **between the ON FREQUENCY sound of a 500Hz CW-filtered transmitter and an unfiltered signal with excessive bandwidth, if he used a 500Hz or narrower filter in his receiver!** As a matter of fact, I normally transmit through a 250Hz filter in my FT1000D, rather than the 2.4kHz SSB filter Yaesu selected. No one listening on frequency, even DX stations copying my signal near noise level, can tell the difference when I select 2.4KHz or 250Hz bandwidth! The only place transmitter filtering makes a difference is up or down the band from my operating frequency.

This is why we can not tell whether a signal has a proper rise and fall time, sharp level transitions, or any other envelope shape problem when we listen to the actual CW tones through a 500Hz filter. Even a very fast rise-time, with a spiked rise and fall, sounds good (and even looks perfect on a scope connected after the receiver's narrow filter)!

Claim's that a certain shape rise and fall produce a "pleasing-sound" are not true at all. First, our ears can't identify a sound only 5mS long, and second...the receiver's CW filter (assuming it is under several hundred Hz BW) reshapes the waveform to a proper rise and fall!

Why is any of this our concern? Why do we have to work on radios, and suffer with clicks? Certainly not because of a cost issue! All the parts are in the radios. It is a simple lack of good design-engineering, most likely driven by a lack of concern by manufacturers for providing rigs with good signal quality.

What Can We Do?

First, we can let manufacturers know it is ***their problem***. Let's ask the ARRL to publish useful reviews with bandwidth pictures showing a spectral display of CW (and SSB) bandwidth. Let's ask them to check for VCO problems, and publish any abnormalities. Let's rate radios as poor, fair, good, or excellent so readers don't have to be EE's to understand what they are buying (and using).

Radios are too expensive, too difficult to work on, and last too long for us to ignore this problem. We need to stop these problems at the design phase, instead of out in the field.

FCC 97.307

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Federal Communications Commission

§97.307 Emission standards.

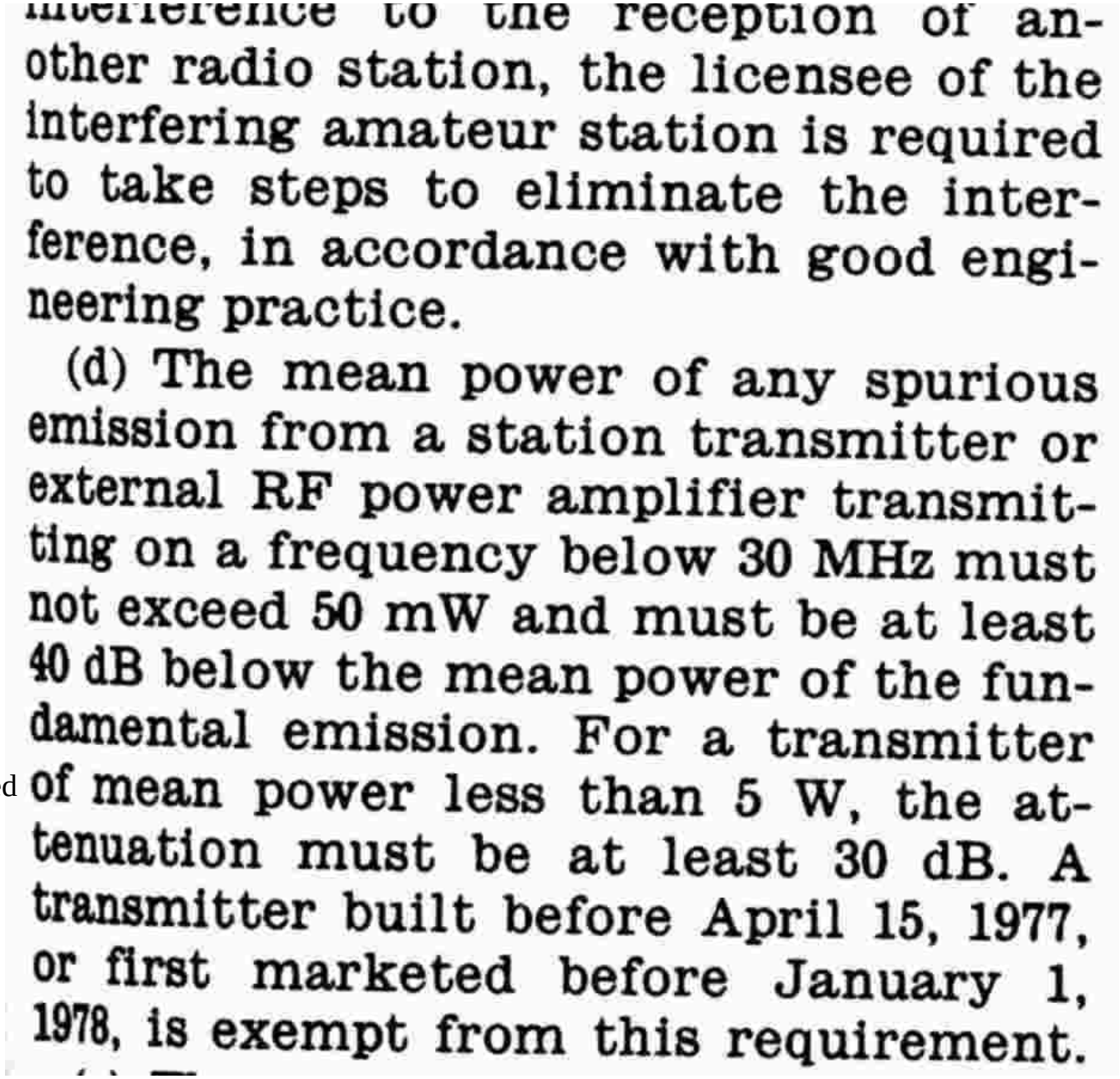
(a) No amateur station transmission shall occupy more bandwidth than necessary for the information rate and emission type being transmitted, in accordance with good amateur practice.

(b) Emissions resulting from modulation must be confined to the band or segment available to the control operator. Emissions outside the necessary bandwidth must not cause splatter or keyclick interference to operations on adjacent frequencies.

(c) All spurious emissions from a station transmitter must be reduced to the greatest extent practicable. If any spurious emission, including chassis or power line radiation, causes harmful interference to the reception of an-

Subpart (a) is very clear. If we only have to change the value of three or four components to reduce bandwidth substantially, do you think it is "good amateur practice" to not do so?

Subpart (b) is also very clear. If the stuff coming out is wider



interference to the reception of another radio station, the licensee of the interfering amateur station is required to take steps to eliminate the interference, in accordance with good engineering practice.

(d) The mean power of any spurious emission from a station transmitter or external RF power amplifier transmitting on a frequency below 30 MHz must not exceed 50 mW and must be at least 40 dB below the mean power of the fundamental emission. For a transmitter of mean power less than 5 W, the attenuation must be at least 30 dB. A transmitter built before April 15, 1977, or first marketed before January 1, 1978, is exempt from this requirement.

than needed, and if you bother someone, you are at fault! Because changing the value of a handful of parts largely corrects this problem, it clearly proves an unmodified radio is illegal to operate near other people.

Subpart (c) is pretty clear also. It doesn't say "when the manufacturer recalls the radios you should get yours fixed."

Subpart (d) doesn't apply to normal modulation bandwidth issues, but if it did we would be in more trouble. The SSB IM3 and IM5 of most radios fails this criteria! Even if a radio did squeak by, as

soon as it was operated into an amplifier it would again become illegal!

I actually had a radio that violated 97.307(d) when used with even a perfect class-A 400 watt amplifier. The radio manufacturer told me "stop using an amplifier with the radio and you'll be fine".

Orion and IC-7800

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Checking bandwidth with receiver

[Home]

Bandwidth rules Part 97.307

Note: ***Bandwidth measurement dynamic range requirements are based on typical signal-to-noise ratios I have observed over the past several years. They are not the extremes of what I have seen, but rather are typical values. Some information on my receiving system and noise floor is available in (NOISE) and (RECEIVING).***

A receiver can be used to check BW if we understand what we are doing! The common mistakes are:

1. **Using a band-scope**
2. **Leaving a noise blanker on (some receivers have a bit of this problem even with the NB off)**
3. **Using excessive (or SSB) bandwidth while tuning to check signal bandwidth**
4. **Using an inferior receiver or a receiver with poor close-spaced strong-signal performance (many DSP-only radios are pretty poor)**
5. **Using excessive gain**
6. **Having excessive background noise or signals**
7. **Relying on an S-meter (most are not linear or accurate).**

1.) Bandscopes

Bandscopes or spectrum analyzers using wide filters can't be used to check signal bandwidth. While such devices are good for monitoring band activity, they have too much bandwidth and are too easily subjected to overload to be useful in determining signal quality or bandwidth. Bandscopes do not have narrow filters and low internal distortion.

You can prove this by looking at a pure unmodulated carrier on the display. The bandwidth of the carrier is nearly zero (it may have some faint noise or jitter), the display should ideally show a perfect single spike of negligible bandwidth.

Bandscopes (and spectrum analyzers with wide filter bandwidths) are all but useless for determining bandwidth or signal defects.

2.) Noise Blankers

Noise blankers must be OFF when checking bandwidth or working close to strong signals.

In order to remove noise, noise blankers add a form of intentional distortion to signals, They do this by turning a switch or gate in the IF amplifiers off and on. An abrupt rise in peak input signal level over average signal level triggers the switch, and shuts the receiver off. The bandwidth of the noise detector is very wide, and this means a strong signal even 5-10 kHz away will activate the NB gate and distort signals.

Some receivers (like Yaesu's) do not fully remove the effects of the NB system, even when the NB is off! In some receivers you have to turn the NB off plus turn the NB gain down, the FT1000MP and FT1000MP MK V are examples of this. In others, like the FT1000D, you actually have to modify internal wiring to correct NB problems. The mechanism is explained in the links to the receiver mods.

3.) Bandwidth

You must select the narrowest filter possible to measure TX BW, certainly less than a 500 Hz filter with good shape factor.

Receiver bandwidth and shape factor directly adds to the transmitter's bandwidth. This means a perfect brick wall 2kHz bandwidth receiver tuning across a perfect 2kHz wide transmitter makes it sound like the actual signal bandwidth is 4kHz. Theoretically it is possible to deduct the receiver bandwidth from apparent measured bandwidth to obtain real bandwidth, but this generally means you have also decreased the dynamic range of the receiver (or spectrum analyzer). In practice, deducting bandwidths often produces unreliable results.

The slope of the receiver (or analyzer) filter is also important. If the receiver response is -6dB at 4kHz and -60dB at 8kHz, you will hear stuff out 8kHz (plus transmitter bandwidth) on very strong signals if you are in a quiet location.

4.) Inferior Receivers

Some radios, in particular DSP only radios, have very poor strong signal performance. They can't be trusted to give accurate BW reports.

Look at tests here, Sherwood Engineering's tests, or ARRL tests of close spaced receiver performance. Many receiver are not all that good. Most Yaesu receivers have a built-in design problem in the noise blanker amplifier that seriously deteriorates close spaced IM performance even

when the noise blanker is OFF.

Even the Sherwood engineering test is too wide for some receivers. The Sherwood test, for example, inflates performance of R4C's with the CF-600/6 filter. This happens because they measured ***outside*** the passband (2kHz) of the filter (600Hz). The second mixer in the R4C is a horrible design, especially the early FET mixer. Close-spaced tests should always be done inside the bandwidth of the roofing filter, or the roofing filter should be considered the narrowest reliably useable selectivity.

5.) Noise

If the band is noisy you really can't check a signal for low-level IM, clicks, or splatter. The noise will cover up any weak signal defects. There must be at least 50dB headroom between the peak signal level being tested and your noise floor to check bandwidth on SSB, or 80dB of signal to noise headroom to check CW bandwidth if you operate near weak signal areas of the band. For general ragchewing away from weak signal areas 50-60dB headroom is generally enough.

Some bands are a special case because SSB operates near weak signal CW stations. 160 meters is one example. Bandwidth of higher power SSB transmitters operated near weak CW stations can be problematic. I often hear spits from Icom 756 and TS 2000 transmitters on SSB as far as 10kHz away on 160 meters.

Receiving noise floor is probably the single most common source of false "clean signal" reports to what actually are problem transmitters. If the noise or QRM floor is high, you won't hear spurious signals.

Wide-audio operators are particular victims to giving each other false assurances of how narrow and clean they are. They often use "opened-up" receivers that absorb more noise power from wide bandwidth (remember noise power is directly proportional to receiver bandwidth) and they often live in noisy environments. It takes a good weak signal narrow receiver in a quiet location to properly check bandwidth.

Results

The result of the factors above is that some people will report a nasty signal "clean" when it isn't, and some will report a signal "wide" when it isn't.

Understanding how to test will correct problems, and help us use our own equipment better. Receivers make very good measurement devices if used ***properly***.

Receiving

[Home] [Up] [Beverage Antenna Construction] [Echelon-Log Beverages]
[K9AY_Flag_Pennant_Ewe] [Magnetic Loop Antennas Receiving] [Receiving_basics]
[Slinky and Loaded Beverages] [Small Vertical Arrays] [W8JI_RX_ANTS]

Noise and common mode noise. Power line and other noise sources. Pre-amplifiers.

This page has top links to other receiving antennas such as Beverage, loop, and vertical antennas. This area deals primarily with low noise antennas, and discusses effect of antenna directivity on weak-signal reception.

Noise Floor and signal Levels at my location:

My local wintertime 350Hz BW noise (after amplifier) compared to a sample of signals on one night was:

Noise -127dBm

9H1BM -122dBm

OM0WR -95dBm

DF2PY -88dBm

WA8OLN -78dBm

W3GH -60dBm

W4ZV -32dBm

The dynamic range between noise and W4ZV was 95dB! It also illustrates how important antennas, location, and propagation are rather than power. The dB difference between signals from the same area can be profound. Many DX signals run at or near noise floor, while others can be 40dB out of noise floor. The data above is certainly not typical of every night, but it shows how large the signal level variations between weak DX and strong signals can be. Over a period

of time I've found 85 to 90 dB dynamic range about the most that is every needed. In a simple installation with a single Beverage, 80 to 85dB IM3 and blocking DR is probably enough. Receivers with less than 80dB IM3 and BDR probably compromise a reasonably good station's capabilities.

Comparison or Ranking of Receiving Arrays or Antennas

The following is a ranking of receiving antennas based on noise being evenly distributed in all directions. These rankings are most accurate in the frequency range of AM broadcast, 160 or 80 meter bands when:

1.) The receiving location shows a nighttime increase in noise level. In other words the system is not limited by local or internally generated noise, instead being limited by skywave or distant propagated noise.

2.) Thunderstorms or other local noise such as power line noise from specific directions does not dominate the receive system noise floor.

There will be occasional exceptions, but as a general rule the ratio of peak response in the direction of the signal to average response in all directions is the dominant factor in determining how good an antenna works for receiving. In virtually all installations without clearly dominant direction or directions of noise arrival, RDF (receiving directivity factor) very accurately predicts receiving antenna performance.

RDF (directivity) will be an almost perfect indicator of what you can expect from your antenna as long as:

- **Noise is not from the same general direction as the desired signal**
- **Noise field strength is not greater than the ratio of peak antenna response to depth of the pattern in the direction of noise**
- **Noise is not coming from within the antenna's nearfield or Fresnel zone**

In the vast majority of systems, the following RDF table indicates relative performance of antennas:

Antenna Type	RDF (dB)	20-degree forward gain (dBi)	Average Gain (dBi)

1/2wl Beverage	4.52	-20.28	-24.8
Vertical Omni, 60 1/4wl radials	5.05	1.9	-3.15
(Ewe Flag) Pennant	7.39	-36.16	-43.55
K9AY	7.7	-26.23	-33.93
1/2wl end-fire Beverages	7.94	-20.5	-28.44
1-wl Beverage	8.64	-14.31	-22.95
two verts optimum phasing 1/8 wl spacing	9.14	-22.46	-31.6
two 1wl Beverages Echelon 1/8 wl stagger	10.21	-15.45	-25.66
Small 4-square 1/4 wl per side (opt. phase)	10.70	-15.79	-26.49
1-1/2 wl Beverage	10.84	-10.88	-21.72
Small 4-square 1/8wl per side (opt. phase)	10.97	-30.28	-41.52
Single 1.75wl Beverage	11.16	-6.50	-17.66
2 Broadside 1.75wl Beverages .2 wl spacing	11.36	-3.51	-14.87
2 Broadside 1.75wl Beverages .4wl spacing	11.91	-3.50	-15.41
.625wl x .125wl spaced BS/EF vertical array	12.5	-19.5	-32.0
2 Broadside 1.75wl Beverages 5/8 wl spacing	12.98	-3.50	-16.48
2 Broadside 1.75wl Beverages .75wl spacing	13.48	-3.49	-16.97

If antennas are within two dB of each other in RDF, a lesser ranked antenna may occasionally outperform slightly higher RDF antennas. This is because:

1. ***Direction and polarization of arriving signals and noise constantly vary, and so the***

relative relationship of each to any individual antenna's response will vary.

2. ***Through various unavoidable errors or omissions, antennas in the real-world may not work precisely as predicted.***

Gain vs. Directivity Myth

One common rumor or myth is that higher antenna gain results in improved reception. Gain is an unreliable way to predict receiving ability on frequencies below upper UHF! A clear example is illustrated above by comparing gain of the single 1.75wl Beverage to the pair of 1.75wl phased Beverages that are spaced .2wl apart.

In this case, the single Beverage has a gain of -6.5dB. The pair of Beverages has a gain of -3.51dB, a gain of about 3 dB. Despite the gain change, antenna directivity and pattern do not change a noticeable amount. RDF only increases 0.2dB, and undetectable difference. Pattern remains essentially the same, so reception remains the same.

Spacing must be at least be 1/2 wl or more for phased Beverages to add noticeable improvement to reception. Even at 3/4 wavelength spacing the directivity improvement falls short of 3dB!

Gain of any spaced pair is about 3dB more than a single Beverage, but reception improves and antenna pattern changes only with relatively wide spacings.

Of nearly equal importance, end-fire arrays actually work better with closer spacing. For an example, compare the 1/8th wl four-square RDF with the 1/4-wl four-square array.

How well does the above hold true?

Over the years, I have had virtually all of the above systems. I always have multiple phase-locked receivers on multiple antennas listening in stereo or a very fast way to "A-B" antennas. When an antenna sits unused most of the time, I replace it with a more useful antenna. My single Beverages are now virtually all eliminated, my last phased loops were in the 80's (when I had four end-fire diamond terminated loops). Even on 80 meters, my large arrays with 300-350 foot spacing almost always beat my single long Beverages. I've migrated towards the bottom end of the chart with all my antennas ***because they actually do receive better.***

If you ask operators who visit for contests, everyone prefers the large vertical or wide-spaced Beverage arrays. Guest operators, given a choice, almost never not use single Beverages or close-spaced Beverages.

You can listen to directivity examples on my DX Sound files page.

Beverage Antenna Construction

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Link to Beverage Antenna Components I use

I installed my first Beverage Antenna (in the early 1970's). I was delighted to find a large improvement in weak-signal reception from such a simple, inexpensive antenna. Over the years I've continued to use, compare, and refine my Beverage antennas. Despite having very large vertical arrays, Beverages remain my primary DX receiving antennas. There just isn't any antenna that is as simple, as easy to construct and maintain, and as foolproof as a Beverage!

I refine my antenna systems by comparing systems against each other for extended periods of time, usually more than a year. My station has a convenient switching system allowing instant comparison of antenna systems. When an antenna system is almost never used, I abandon that system and try something else. Even though I use engineering tools (books and models), I always compare and measure actual working systems. I presently have over thirty Beverages in three different clusters of arrays, the end result filtered through years of measurements and A-B testing of systems.

A great much has been written about Beverages. Unfortunately much or most information is a repeat of previously published information (and misinformation) from verbal discussions or from other articles or handbooks! It is time to set aside some of the myths that have been handed down and repeated so much they have become "fact".

Types of Beverage Wire

The most commonly used wire types are single conductor hook-up or electrical, electric fence wire, and special antenna wire such as copperweld. The only significant and easily noticed difference between these commonly used wires is in physical properties, such as ease of soldering, strength, and life

Insulated Wire

Sporadic claims have appeared indicating insulation prevents charged droplets of water from making an antenna "noisy". I've never been able to verify that rumor either in A-B tests of actual antennas or through planned experiments. Other reports, many from reliable sources, also seem to discredit this rumor.

One of my experiments was to charge a stream of water (against earth) with an extremely high voltage supply, and spray the water on a wire. Other than corona noise from sharp points, the type of wire made no difference at all in "noise". The water droplets obviously were not hitting the wire like hundreds of random charged capacitors, they generated no noise at all. This is really what we would expect, if we consider that each drop contains only a very miniscule amount of change and also has nearly perfect insulation (distilled water is a very good insulator).

Controlled observations also tend to support the idea corona, and not charges in individual droplets, cause precipitation static.

In Ohio, my long Beverages stretched across open farm fields. Snow would whip across the fields, rain would pelt the wires, yet insulated and bare wire Beverages running in the same direction always had the same noise level. Beverages that picked-up corona (or "p-static") noise were always near or aimed at tall towers. With corona sizzling at 40-over-nine on my tall towers, Beverages (and even small "magnetic" loop antennas) aimed at the towers would "hear" the same precipitation noise.

The same was true for tower-mounted antennas. The largest noise problems came from antennas mounted high on towers, and generally were with antennas that had "sharp" ends jutting out in the air. Lower antennas, even those of identical construction, were either significantly quieter or totally free of precipitation static. This effect was reported many times by contest operators and DX'ers with stacked antennas. They universally switch to low antennas to eliminate or reduce p-static, even though the ***same moisture is hitting the lower and upper antennas.*** This strongly indicates precipitation static is from corona discharge, and ***not*** from charges in each individual drop of moisture hitting the antenna.

After my move to Barnesville, Georgia my first antennas were all insulated wire. Hook-up wire was pressed into service in my first group of temporary Beverages. As non-insulated conductors permanent antennas were added, there wasn't any observable change in inclement weather noise. As before, the antennas nearest or aimed at tall towers picked up some p-static noise. Antennas located away from the towers remained free of precipitation static, whether bare or insulated.

There is also some chance, if the antenna wire is not under significant tension, that insulation may sometimes hide a broken conductor.

Insulated wire may reduce leakage currents if a substantial part of the conductor is in contact with resistive paths, such as wet brush or tree branches, but you may be better off trimming back any substantial foliage in contact with the wire.

While insulated wire has no major performance disadvantage, it also has no advantage.

Type of Conductor

Copper wire is a good choice if supports are close. Copper wire lacks the mechanical strength of steel-core wires, but is very easy to work with. It is softer, making it easier to bend. Copper wire can be repeatedly scraped or re-soldered without worries about piercing a thin copper coating and exposing a rust-sensitive steel core, and it is readily available and relatively inexpensive in large quantities.

Copperweld wire is much stronger and has about the same RF resistance as 100% copper. Like copper, it is easy to clean and solder after it has been exposed to the weather as long as you are very careful to not scrape through the thin outer coating of copper. It is considerably more difficult to work with than normal pure copper wires, any kink or sharp bend will substantially weaken the wire.

Most fence wire I've found is cadmium plated, rather than zinc galvanized. Using RF current meters, I have measured increased losses when using zinc or cadmium plated steel wire. Beverages already have substantial current loss due to the close proximity with lossy earth. I've measured about 60% of feed point current remaining (~4.4dB loss) after passing over around 700-feet of electric fence wire, and about 10% more current (~3.1dB loss) using copper-clad steel wire. Steel fence wire would aggravate losses that already limit the benefits of using long Beverage antennas. *In a very long antenna, the small*

additional loss of steel fence wire might slightly reduce performance.

In my Beverages, the important consideration is antenna maintenance. I use copperweld wire or electric fence wire, because strength is a primary concern. With spans exceeding 200 feet, my antennas need a large strength to weight ratio.

Don't use welding wire! It is a very poor material choice. It rusts (and as with aluminum) you'll have connection problems in no time.

Beverage Supports

Some would have us believe we need non-metallic supports for our Beverages, but there is not the slightest technical justification for such suggestions.

The only requirement for the support is it must hold the antenna up, and it can not connect the antenna to ground. A metal pole with a small PVC stub for an insulator is every bit as good as a full non-metallic pole. Trees make good supports, especially if you use nail-type electric-fence insulators for use with wooden posts.

I've never seen a problem allowing a wire to contact a branch, although I do trim out the branches and avoid any contact with trees.

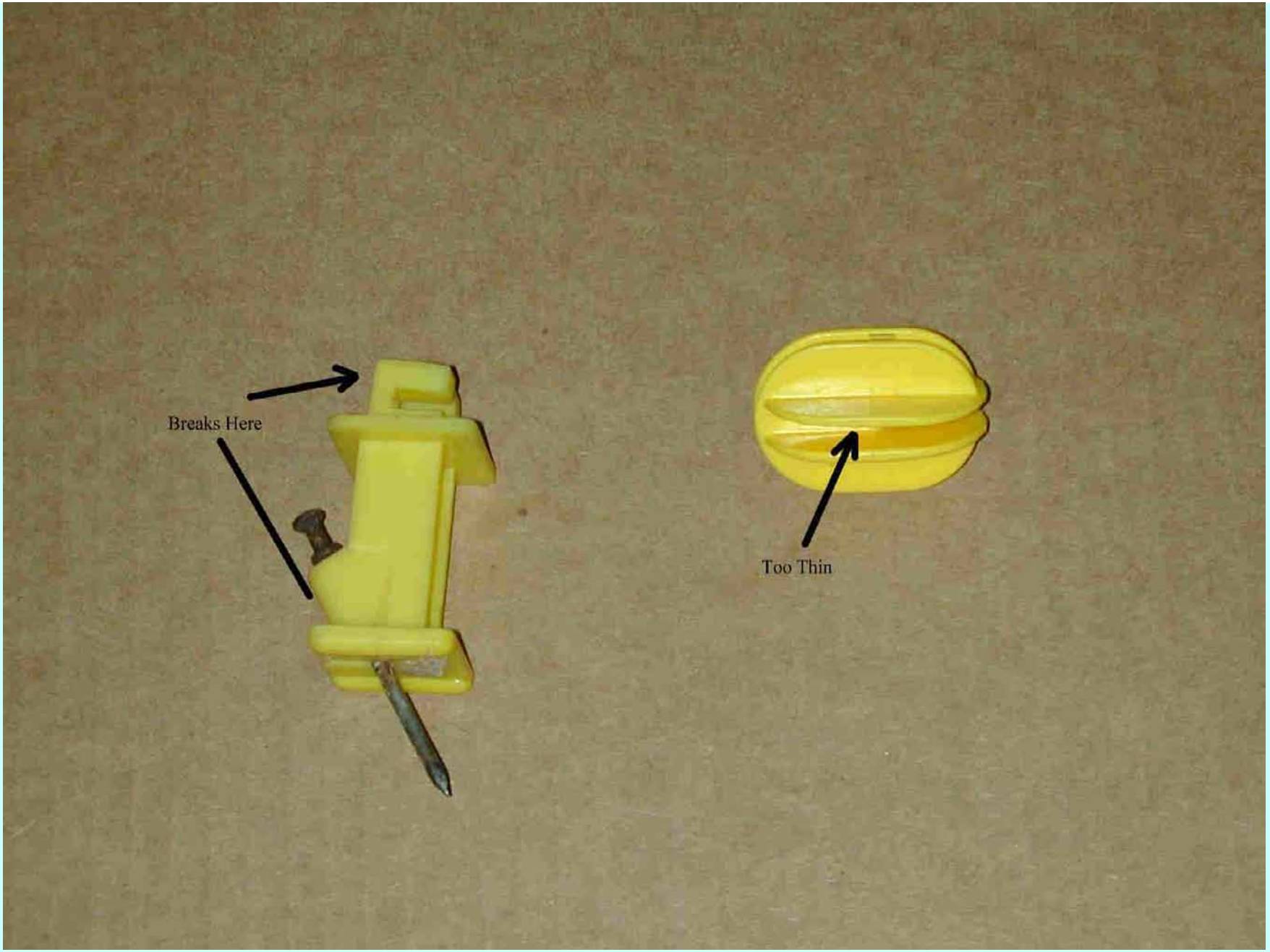
For end supports, I use trees, pressure treated lumber, or landscaping timbers. With a lot of tension, I backstay the poles to a dead-man (generally an old brick) buried in the ground. When I set end-posts with my power auger, I line the hole with copper flashing. That becomes part or all of the feed point (and termination) ground connection.

I never anchor or wrap the Beverage wire around insulators, except at the ends. I always allow the wire to "float" through the insulators. When the wire floats, you can tension the entire antenna from either end. If anything breaks the wire, you can see it at any point! A "floating" wire is much easier to repair if it is damaged, because you only need release tension on one end to splice the wire. Re-tension that same end, and everything is restored. It takes no more tension to support a 1000-foot Beverage with supports every 100-feet than it does to support a 100-foot wire between two rigid supports, but it is a much more difficult to break the longer wire. A longer "floating" wire will often take-up enough slack to remain up after deflecting a large tree branch, where a shorter rigidly-anchored span will almost certainly break either the insulators or wire.

Beverage Insulators

If you expect a long-lasting antenna and have a long antenna, be careful when choosing insulators! Some types of electric fence insulators will not last long. The unreliable types of post insulators have two square folds to hold the wire, a square shaped base, and nail through a small molded plastic angle. The weak points of this insulator are the square retaining tabs, and the molded nail tube at the insulator base. When this type of insulator is mounted horizontally, the wire's weight will stress both the molded nail tube and a single tab. I typically find about 10% of the insulators fail within a few months. After three years, the few dozen installed here have virtually all failed.

Avoid these types!



Round yellow or back plastic insulators with the nail going through the center, like the examples below, are much more reliable post insulators

Ceramic post insulators may look great, but they do not allow floating the wire across the insulator. Even if you do manage to find a ceramic insulator that allows floating the wire, the ceramic will quickly wear away at the constantly moving wire. Avoid ceramic insulators, unless you are prepared to "buffer" the wire through a UV resistant soft plastic bushing!

Good end-insulators are becoming difficult to find. I always use compression types, but the material has to be either ceramic or very thick plastic. Some very thin plastic compression insulators will actually cold-flow and allow the wire to pass through the insulation. This is particularly true with thin steel wires that are tensioned over 25 pounds. Heavy-walled egg insulators are much more reliable, and not subject to wire migration through the thin insulation.

My favorite insulators are large these rather thick Fi-Shock yellow plastic insulators. They are slippery enough to allow the dead-end wire or rope to loop over the insulator, and create a 2:1 mechanical advantage when tensioning.

Height

I've found very little performance difference with height, unless the Beverage is more than .05WL high. As the height exceeds .05wl, performance seems to be reduced. Small rolling hills or ravines also seem to make any difference. Follow the contour of gradual slopes, and go straight across ditches or narrow ravines without following the contour.

Sloping Ends

There really isn't a logical reason to slope the ends of a Beverage. After all, six-feet of vertical drop is six feet, no matter if the drop is over 50 feet or straight vertical.

Consider, for example, the K9AY or Pennant antennas. Both have sloped wires, yet virtually all of the response is from the vertical slope of the wire in the antennas. As a matter of fact, the actual shape makes very little difference in the way each antenna works. Why would anyone, knowing how a Pennant or K9AY works, think that a Beverage somehow magically breaks tradition and stops responding to vertical signals in the wires when we slope them a bit? What difference would it make in noise anyway, since the entire antenna responds to vertically polarized signals?

There isn't any possible way, including use of shielding or additional conductors, to prevent the end-wires from having the very small effect they have. Save yourself time and worry, and avoid a needless hazard. Just drop the end-wires vertically right down to earth.

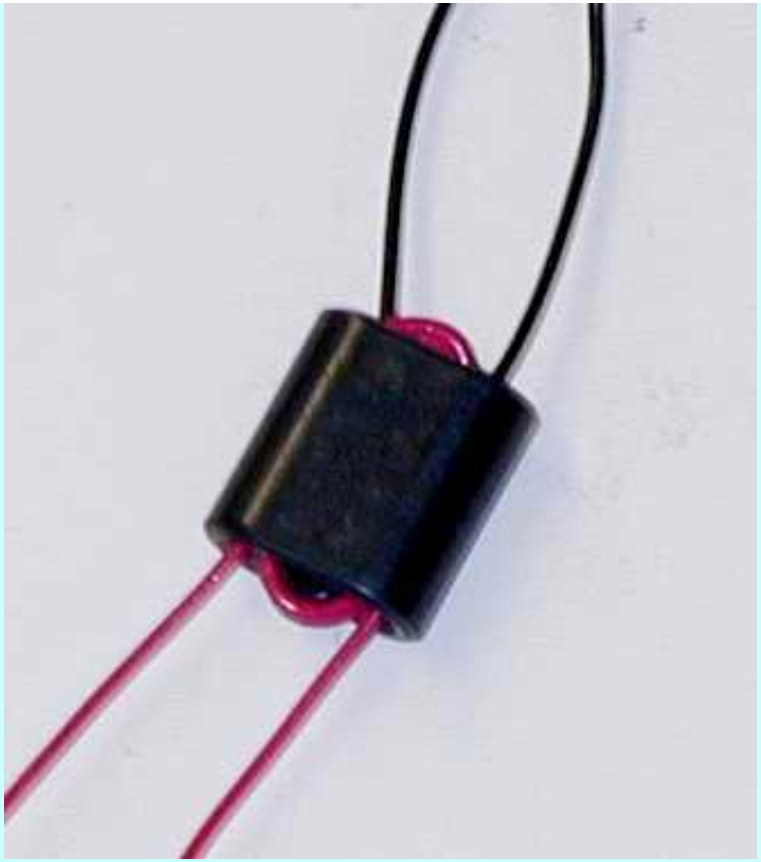
Multiple Antennas Crossing

Crossing of Beverages has little effect if they are not parallel or nearly parallel. Try to cross at an angle of 90 degrees if possible. Even a few inches of spacing is enough for right angle crossing. With shallow angles, assuming they can not be avoided, increase wire spacing to a few feet.

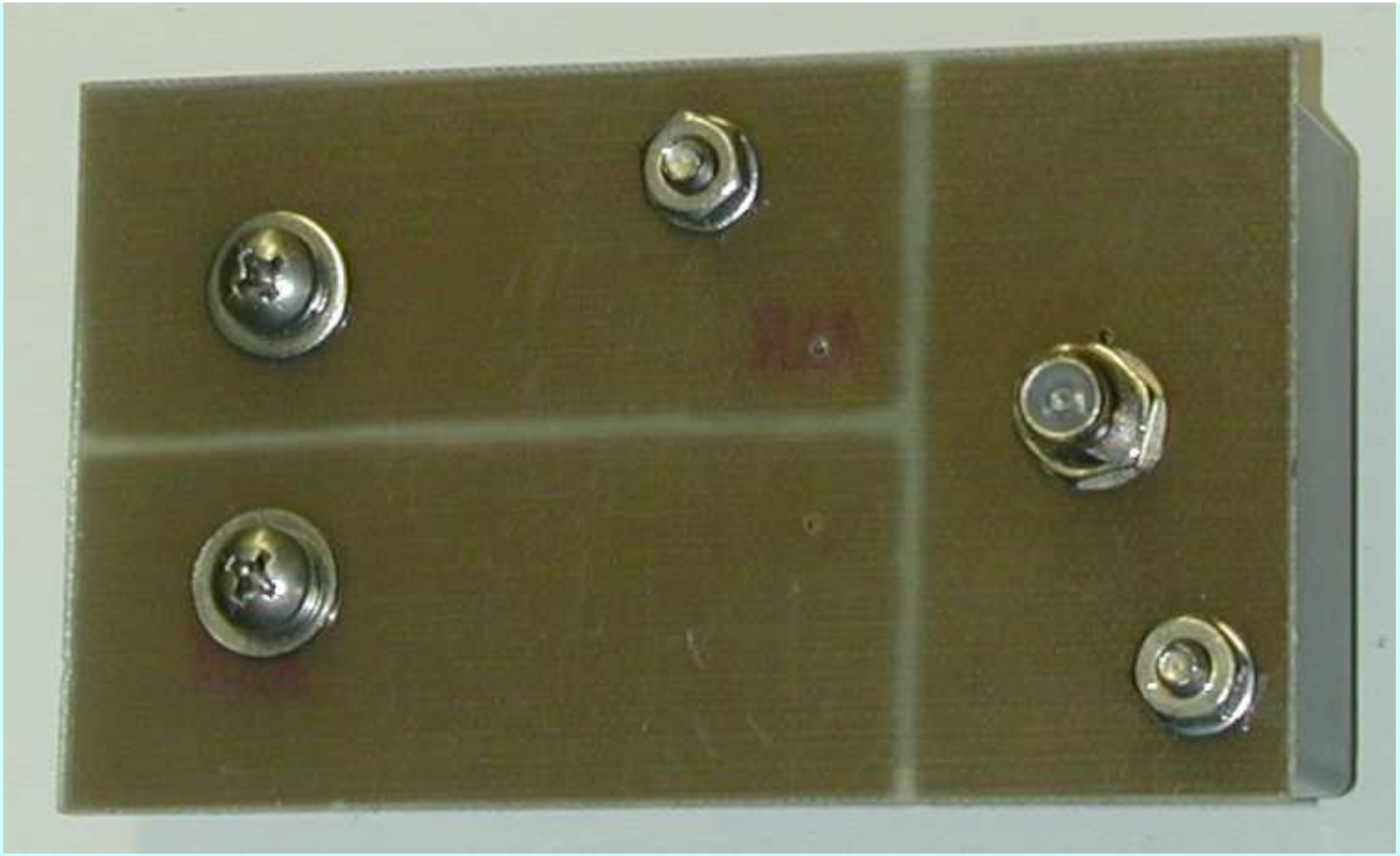
Transformers

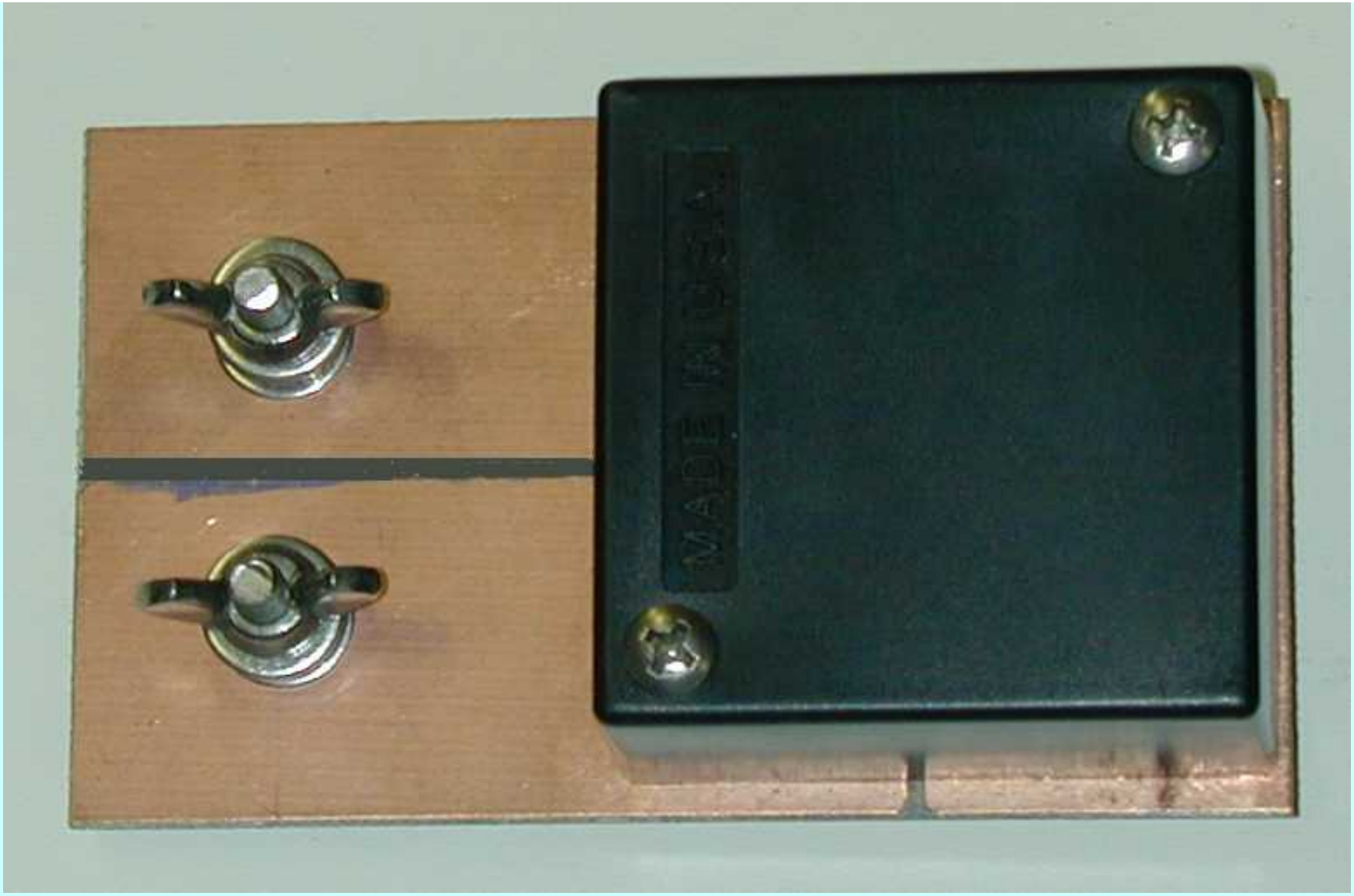
Always use isolated transformers for feeding Beverages. It is cheap, simple, easy insurance against unwanted common-mode ingress of noise and signals into the antenna from the feedline shield. See the Common Mode Noise page for an analysis.

I use 73-mix FairRite Products 2873000202 cores (about 1/2 inch square and 1/3 inch thick 73 material) in my transformers. These cores require a ***two-turn 50-75 ohm winding.*** The high-impedance winding is ***5 turns for 75-ohm cables*** (6.25:1 Z ratio) or ***6 turns for 50-ohm cables*** (9:1 Z ratio). Small insulated hookup wire is actually better than enameled wire. The thicker insulation is much less susceptible to developing shorted turns in rough service.



While my early transformers were waterproofed with Krylon and coated with insulation foam, I have finally laid out enclosed transformers and terminations with internal lightning protection. The transformer sections have F-fittings, and all use stainless steel hardware.





For a Reversible Beverage, I use the following:

Multiple Antennas at One Feedpoint

Never bring multiple antennas to one feedpoint, especially when they share one common ground. I've noticed a definite deterioration in pattern with multiple feedpoints arranged with only ten feet of spacing, even when they had separate ground systems. One set of Beverages installed with 5-10 foot of feedpoint separation has noticeably poorer patterns than other identical length antennas with wide separation at the feedpoint.

Multiple antennas actually may be the only case where a sloped feeder can make a difference, the slope will actually move the effective feedpoints further apart. The best idea, however, is to separate the feedpoints by several times the antenna height.

Termination Value

Having precise termination values isn't necessary, but get as close as you reasonably can. There are some impedance measurement suggestions circulating that absolutely do not work. One is to just use a tuner to match the terminated (or unterminated) antenna, and work backwards with loads to measure tuner impedance ratio after matching. This won't tell you a thing about proper termination, unless you repeat the measurements on dozens of frequencies spread over a wide range!

There are three fast, simple ways to test for proper termination:

With an Antenna SWR Analyzer

1. Connect the antenna analyzer at the Beverage feedpoint through a good matching transformer
2. Sweep the analyzer frequency from 1.8 to 7 MHz (or over a ~4:1 frequency range near the frequency intended for antenna) while watching SWR
3. Adjust termination for minimum SWR ***variation*** (not minimum SWR, minimum SWR variation!)

When installation (including grounds) and termination is proper, ***SWR VALUE*** will remain nearly the same regardless of frequency

With an Antenna Impedance Meter

1. Measure the feedpoint impedance (right at the feedpoint) of a roughly terminated antenna at the frequencies of ***highest and lowest*** resistive impedance. You can do this through a known good transformer by correcting impedance for use of the transformer
2. Multiply the lowest measured impedance by the highest, and then find the square root of that number. This will be the correct termination impedance of the antenna

With a Clamp-on RF Current Meter

(This does not work well with voltage, because of measurement method error problems)

1. Apply a small amount of power from a transmitter, do not exceed antenna system component ratings!
2. Measure current at the termination, and several points up to a distance of at least 1/2 wl from the termination
3. Adjust termination resistance so current shows a smooth current decline as you move the meter towards the termination

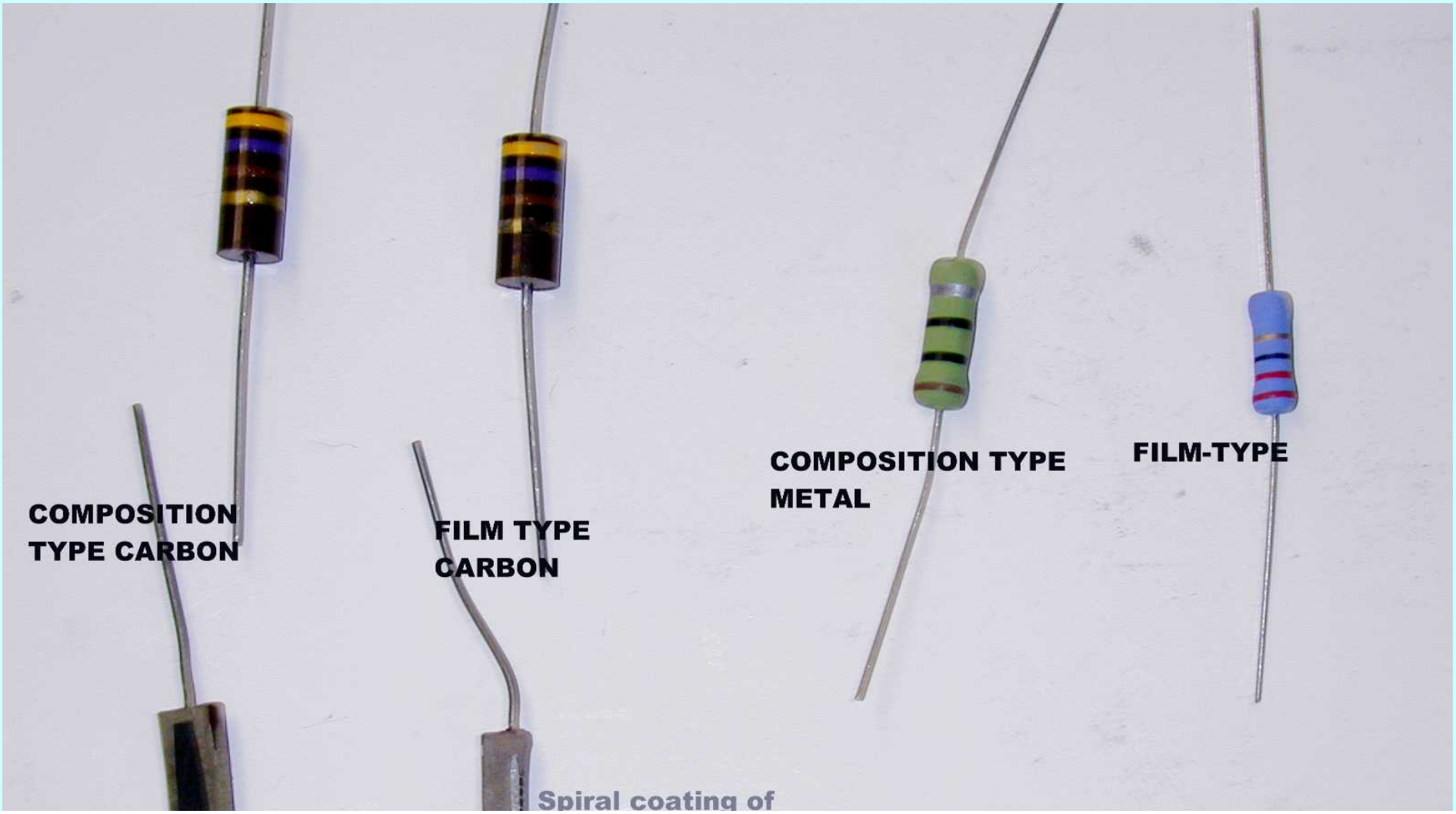
Note:

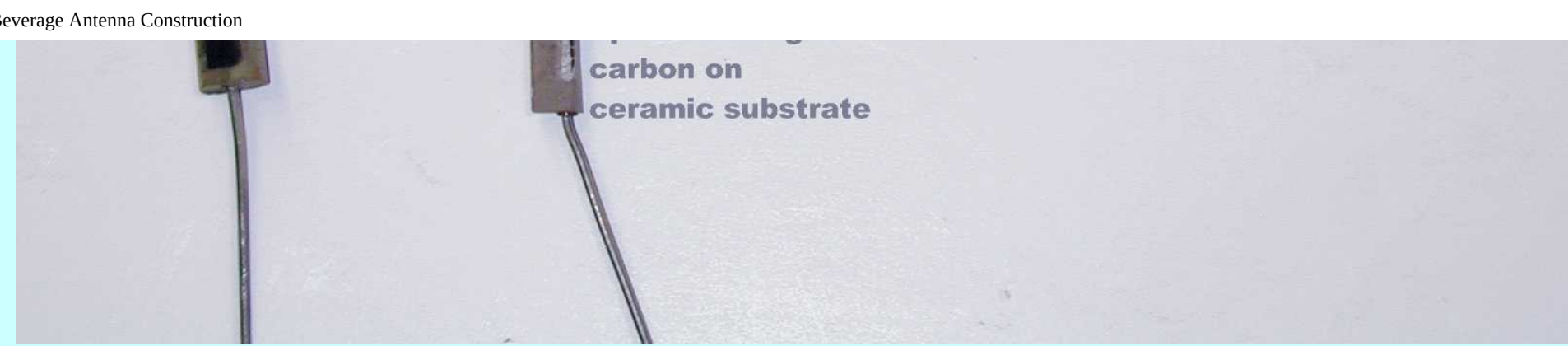
In about 500-800 feet of distance, power loss in a Beverage is around 3dB. This corresponds to a 1/3 reduction in current. If you attempt to adjust for equal currents (or voltages) over any distance, the antenna will be MIS-terminated!

Termination Components

Identifying a Composition Resistor

We commonly assume any brown phenolic resistor is a carbon composition resistor, but that isn't true. Most of the smooth brown-colored phenolic cased resistors manufactured after 1960-1970 have actually been carbon film resistors. There are only a limited number of manufacturers supplying carbon composition resistors. One is Allen-Bradley. They are expensive special-order parts, and the buyer must specify composition types.





As we see from the photo, it is impossible to identify a composition resistor by external appearance.

The only sure way to identify a resistor, short of ordering it from a reputable source, is through a destructive test. We can, for example, apply a large momentary overload and look for a resistance change. A resistance change indicates a film-type element. We could also cut the resistor open, and look for a non-conductive core. A non-conductive core indicates the resistor is a film style component.

Why Composition Types?

We need composition resistors in any application where the resistor is subjected to very-large very-short overloads, or where the system demands a nearly pure resistance at a very high frequency (F>100MHz).

Obviously, in the case of a Beverage at a few MHz or lower, we could get away with using many styles of wire-wound resistors or spiral-film resistors. A small amount of inductance would not be a major problem, and virtually ALL carbon or metal film resistors (constructed with resistance elements deposited or cut in a spiral on an insulated core) would not have excessive inductance. The thing we can not tolerate is the sensitivity of non-surge rated components to damage from lightning storms, even distant storms.

The life of a carbon or metal film resistor, when used as an antenna termination, is relatively short in most locations. Just a few coulombs of energy, when applied in a few milliseconds, will cause a carbon or metal film resistor to change value. Worse yet, the resistor will not be altered in appearance. (Carbon also has a strong tendency to change value with heat. Even modest operating temperatures, over a period of time, will cause a carbon resistor to change value. Metal resistors are more stable.)

Unless you want to make a full-time career out of testing your antennas and replacing resistors, use a energy absorbing composition type resistor!

I install a small lightning gap of about 1/8th inch across my antenna's ends, both at the feedpoint and the termination. This helps immensely with very close strikes. I use either Ohmite OY-series metal compositions or A-B carbon composition resistors. You can buy metal composition resistors at DX Engineering.

Ground Systems

The ground system mainly provides an RF and lightning ground. Having a very low ground-resistance is not especially important, unless an Autotransformer or Un-un is used! Autotransformers and Un-un's don't isolate the feedline for common-mode. The antenna needs a stable ground, not necessarily a low-resistance ground.

In my tests over the years, a 3/4-inch copper pipe driven five feet or deeper into the soil typically measures between 50-150 ohms of RF resistance on 160-meters. (DC or low frequency AC measurements will NEVER give the correct earth resistance for RF, and they certainly can not tell us ground conductivity.) Unless you have exceptionally poor soil, going deeper than five feet will not reduce RF resistance on frequencies above 1.8 MHz. Skin effect limits the depth of RF current in the soil, so the extra rod depth does nothing. Lower resistance values (about 55 ohms) were obtained in a wet marshy area of NW Ohio, with a very rich black acidic sandy loam soil. The higher resistance were obtained in rocky clay soil typical of the Atlanta, Georgia area.

My present location has rolling pastures and wet clay soils, providing under 100-ohms of RF resistance at 1.8MHz with a five-foot rod.

The general guideline I follow is to use at least two five-foot copper rods (I use 3/4" copper spaced 5 feet apart). If I can not get full depth, or if the soil is particularly poor, I add a few 30-60 foot buried radials. The idea is to obtain a reasonably stable ground, so termination does not change.

CLICK TO LOOK AT ACTUAL MEASURED GROUND TERMINATION RESISTANCES!!!

If you are unsure if you Beverage's ground is adequate, measure the impedance of the beverage with an antenna analyzer with your operating ground systems. Note the reading. Add two temporary radials 1/4 wl long suspended above earth at right angles to the Beverage, and re-measure the impedance. (It is OK to have them there at right angles to the antenna and not have them connected, and them connect them while taking readings.)

You can measure the impedance on the low-Z side of a good transformer. Under almost any condition, the wires would have 100 ohms or less impedance. If you see a very noticeable change in impedance, you probably should consider improving the ground system. Impedance changes of 15% (or larger) indicate a potential ground stability problem, because the ground resistance would be nearly 100 ohms. This test should be done when the ground is dry, or any time you think you might be having a ground problem.

Always remember to keep the shield of the cable isolated from the Beverage ground! Never use un-un or autotransformers.

Length

For length considerations, see the directivity factor text. It is not necessary, nor does it do any good, to go beyond 1-1/2 or 2 WL. By the time the antenna is that long, current is so low any addition length makes the pattern worse. I limit my 160-meter antennas to 800-feet, and use multiple antennas when a sharper pattern is required.

Directivity can actually decrease if a longwire-type array is made too long. This is true with Rhombics and Vee Beams, and it is also true with Beverages.

Zigzagging Wire

While a nice clear straight wire looks great, it does more to make us feel better than hear better! Minor ups and downs in height or dips or valleys don't really seem to have any noticeable impact.

Although it probably is a good idea to keep the wire as straight as possible, it is the overall direction and length that is most important because each small area contributes on a similar small portion to the overall directivity and signal reception.

Verticals and Baluns

[Home]

Related articles at

Balun Test contains model of "perfect" dipole currents.

Sleeve Balun shows how a sleeve adds impedance, useful for VHF and higher baluns

Receiving Common Mode Noise shows how lack of a balun can contribute to system noise (it applies to transmitting antennas as well)

Balun and Core selection for transformers and baluns

Transmitting baluns on testing transmitting baluns

RF in the Shack

Many people assume that RF in the shack or (worse yet) RF burns are tied to problems from poor station grounds. With properly operating coaxial feedlines or balanced feedlines, the operating position should have minimal RF even absent a shack ground. Only bringing a longwire or some other single wire feeder directly into the shack should cause RF burns or RF feedback in the shack.

There is one exception to the above. Direct radiation from the antenna into station wiring could cause high levels of RF to appear on equipment or wiring, but even in that case the station ground is not the problem or cure!

Contrary to popular belief that only dipoles need baluns, verticals and longwires can require baluns (more correctly called common-mode chokes or isolators in this application) also. This article shows why baluns (or common mode chokes) might be required, and what the balun can do.

What causes common-mode current?

Current flows because there is a voltage difference between two parts of a system along with a return path for the current flow. The path can be "special currents" without actual electrons flowing called

"displacement currents", or it can be actual charges moving through conductors. Displacement currents flow through the dielectric of a capacitor, between a vertical or single wire fed antenna and the "ground" for that antenna, or currents that flow between a mobile's antenna and a car body.

Displacement currents commonly complete the entire current path in antennas. They are the sole reason current in physically large coils can vary from end-to-end, and are especially problematic in mobile antenna installations. They are the reason an open-ended antenna like a dipole, longwire, or vertical is able to have current flow!

When we force charges up into a Marconi vertical or longwire antenna (making current flow), we have to move an equal number of charges out of some ground system or counterpoise into the feedpoint. The ground can be a single conductor or many dozens of wires, and it can (and often does) involve equipment in the house and/or the coaxial feedline shield. The bottom line is we always must have the same current coming back to the feedpoint as moves up into the antenna! There is no way around that rule.

This creates two problems:

1. **We may not be able to handle all the charge displacement with a few radials without having significant voltage driving those radials**
2. **The feedline has to connect to the antenna, the outside of the shield may be excited by this voltage, and become part of the return path for "collecting" displacement current**

Another way to view this is the feedline has to have something to ***push against*** to force current into the antenna. It is like pushing a car. If you have very poor footing, your feet will move and slide. The same is true for a ground system, as feedline power "forces" current up into the antenna the other terminal of the feedline has to be held steady.

The Feedline

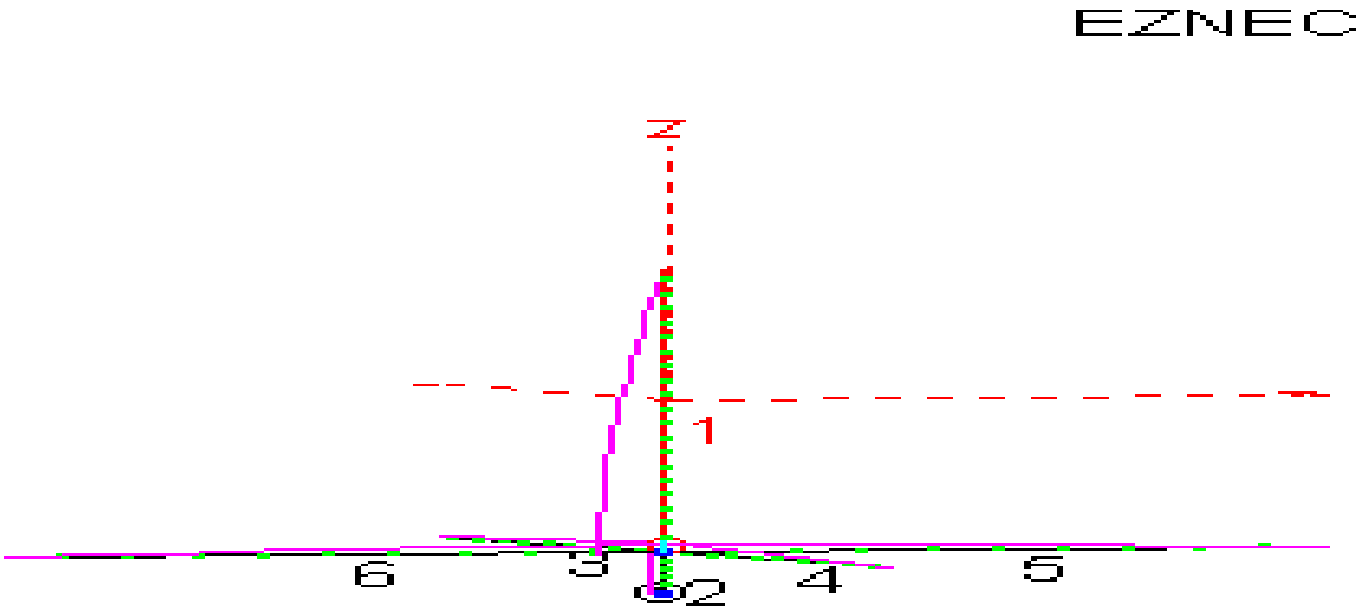
If you are not familiar with how coaxial cables work, you might want to look at a simple explanation on this site or one of the ARRL Handbooks.

In order for a conductor like the outside of the shield to ***not*** have current flow at Radio Frequencies, it must have the same electrical potential and phase all along the length. If it has a high series impedance or if the potential difference along the conductor is low, very little current will flow. As seen in coaxial cable operational descriptions, any coaxial feedline can have unwanted common mode currents.

Does a vertical or longwire present common mode voltages to the feeder that can cause common mode currents? You bet! The only vertical (or longwire) that would not cause such problems is one with a very good or nearly perfect ground system, and that means something that looks like a large infinite groundplane. Even then, the cable must exit ***below*** that groundplane to be "shield current free".

Aren't Four Elevated Radials Perfect?

Here is a model of a groundplane with four radials:



EZNEC ver. 3.0
Balun 80 vertical 1/3/04 7:19:05 PM
---------- CURRENT DATA ----------
Frequency = 3.6 MHz.
Wire No. 1: 6.700
Wire No. 2: 1.359 (This is your feedline or mast)
Wire No. 3: 1.985 (These are the radials)
Wire No. 4: 1.985
Wire No. 5: 1.985
Wire No. 6: 1.985

We can see significant current flows over wire 2, which would be the coax shield, a mast, or both.

There is a trick with Eznec. By inserting an additional source in the mast or feedline and setting current to zero, we can observe the radial common point to earth voltage required across a balun to force current to zero. In this case the voltage across the balun would be:

Source 2 Voltage = 145.5 V. at 67.97 deg.
Current = 0 A. at 0.0 deg.
Impedance is infinite

Amazing isn't it? At 1500 watts the ground common point for the radials actually wants to have ***145.5 volts to earth*** to prevent current flow and unbalance of the system!!! If we elevate the common point to 145.5 volts at 68 degrees phase angle, we now have the following currents at 1500 watts:

Wire No. 1: 6.4 A
Wire No. 2: 0 A (coax shield or mast)
Radials: 1.58 A each

Other Systems

With fewer radials the situation becomes much worse! (As a matter of fact, this is a good reason to use as many radials as we can even if the radials are resonant.)

Claims that four elevated radials form a "perfect ground better than 120 radials" are obvious nonsense! If it was a perfect ground, there would be no potential difference to earth and no common mode current flowing to "real ground" !

If the antenna has a high base impedance, it will have less current at the feed connection. (Sorry, but we cannot do this by using a folded unipole!)

Longwire and Windom antennas really aren't much different than verticals. They are a form of Marconi antennas, and require a counterpoise or ground of some sort. As with verticals, common mode current flowing into the antenna must be balanced by current flowing into a ground system.

Instead of bringing the longwire directly to an antenna tuner, a better solution is using an RF ground system independent of the station safety ground, and keeping that ground isolated from the station safety ground. That can be accomplished by adding a good heavy-duty 1:1 choke or current balun a few feet from the tuner, and connecting the RF ground to one output terminal and the antenna to the other terminal. With a two-wire feed Windom (really an OCF dipole), the two wire should connect to the choke balun.

The balun MUST be a current balun, rather than a voltage balun.

Summary

The cure for common mode problems caused by less-than-perfect grounds is inserting a 1:1 choke balun in the system at the antenna feedpoint. The coax should also be kept away from the radials as it exits the area of the radials and the antenna. An antenna with a poor ground using few radials cannot have a support mast grounded to the radial common point (at least it shouldn't if designed properly). There is no exception to this!

For comparisons of baluns see balun test.

RF in shack with vertical and longwire antennas

Transmitting Baluns

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Related articles at

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Testing Transmitting Baluns

One of the most popular Ham-lore rumors is a balun's performance can be tested or evaluated by grabbing the coax and watching for an SWR change. This is probably one of the worse test-rumors circulating!

An observable SWR change when touching, grounding, or altering length of a feedline (with the same Zo at the test meter) indicates severe common-mode current problems.

The opposite response, however, means nothing. Lack of SWR change does NOT prove the balun adequate and the system free from common-mode currents.

I was first exposed to this wild idea when asked to evaluate a problem with a commercially manufactured antenna. The antenna's design engineer decided a bead balun was adequate, based on this "grab and touch test". The antenna/balun combination passed the "grab test" with flying colors, even though the system had gross common-mode current levels. Despite extending the bead balun from the original foot to almost three-feet of beads, I couldn't remove common mode excitation of

the feedline by the grossly unbalanced antenna.

If we think about it, we would never consider it likely that our hand would have significantly lower impedance than the shield on a long length of coaxial cable! We also would never expect, just by dumb luck, to always grab the high impedance point of a random wire like the feedline.

A hand-grab-test will cause an SWR change only when the impedance of your hand is very low compared to the common-mode impedance of the cable at the point where you are "grabbing" the cable. The portion of impedance attributed to feedline radiation (compared to overall antenna impedance) must be a significant portion of antenna terminal impedance or the SWR will not change.

It is possible, if we alter antenna feed-cable lengths, to observe SWR changes as an indicator of common mode currents from poor system design. But such methods are never nearly as accurate as an absolute current measurement, and are actually only a little bit better than the useless "hand test". They also can give false answers if the feedline impedance is not the same as the SWR measurement device impedance.

An Accurate Test

The most simple and accurate way to test the effectiveness of a balun is to actually measure the common mode currents on the feedline with a suitable RF current meter that does not perturb the system. Such meters are inexpensive to purchase and are also easily manufactured at home. The meter would consist of a closed core that snaps around the cable, a terminated winding on that core, and a meter that measures RF voltage across that winding. The coil's load resistor should be low enough in value that the impedance of the cable shield is not perturbed.

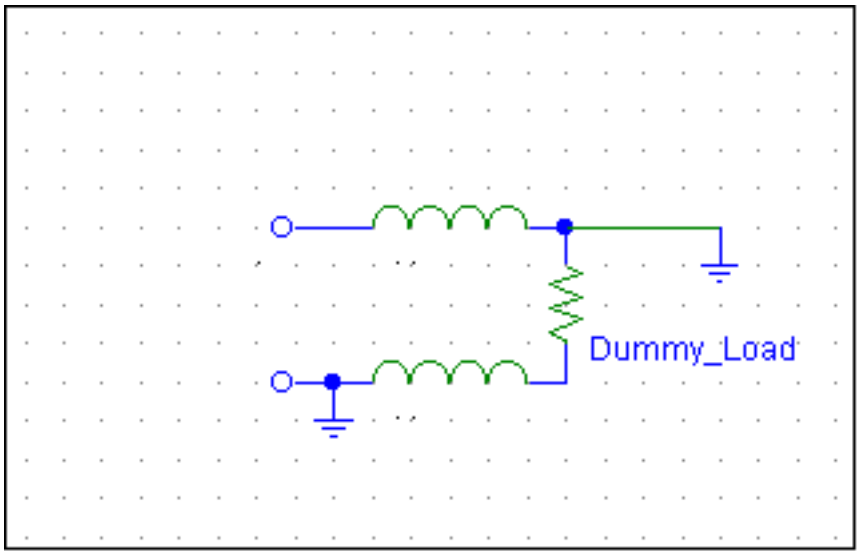
By sliding the meter along the feedline, we could get a good idea of the common mode currents. The meter should be moved along at least 1/4wl of feedline. It normally would be sufficient to sample just four points along that distance.

Feedline radiation, like antenna radiation, is a function of the linear ampere-feet of conductor. The important distance is the spatial or linear distance, not the length of a coiled conductor. It is the charge acceleration over a linear distance of space that accounts for the radiation. A one foot box packed with 500 feet of cable is still just a one-foot antenna!

There are many cases where adding a balun will actually INCREASE common mode currents. Be sure you either fully understand the system or have modeled the feedline and antenna correctly.

Power Testing

A choke or current balun can be tested by inverting the ground on either end, and using a dummy load. Actual operation may produce more stress than this test, or less stress, but such a test gives a reasonable idea of the power rating under reasonable conditions.



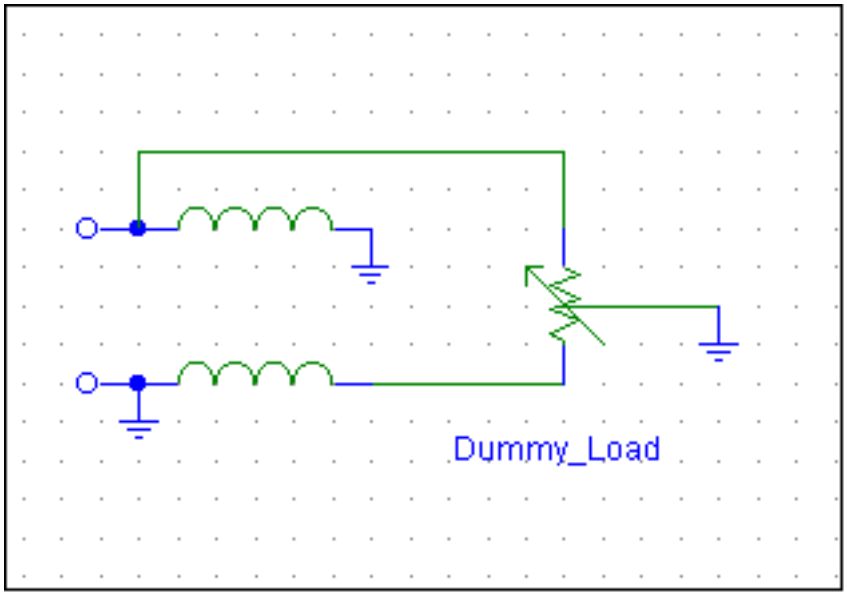
The balun should not overheat or change impedance with application of full power for the expected duty cycle in the expected maximum operating temperature (be sure to include sunlight heating).

Impedance Testing

Using the above test configuration, SWR can be measured. SWR should be flat, and should not change when the ground by the dummy load is moved from side-to-side. This should be true at full sustained operating power.

Testing Voltage Baluns

While voltage baluns are the least desirable types in most applications, they still have some useful applications. Using this configuration:



By moving the tap point along the dummy resistance, an idea of balance can be obtained. The balun should show perfect performance with the tap-point at the load center, and should show an SWR increase as the tap moves to the top or bottom.

Crossed Split Windings on Toroids

Some articles and handbooks show a Split winding method. This method is supposed to reduce winding capacitance by moving the ends of windings further apart. The proposed theory is by reducing shunt capacitance that "leaks RF around the balun", balun performance is enhanced.

This method very often hurts the performance! Check this article

toroid balun winding

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Related articles at

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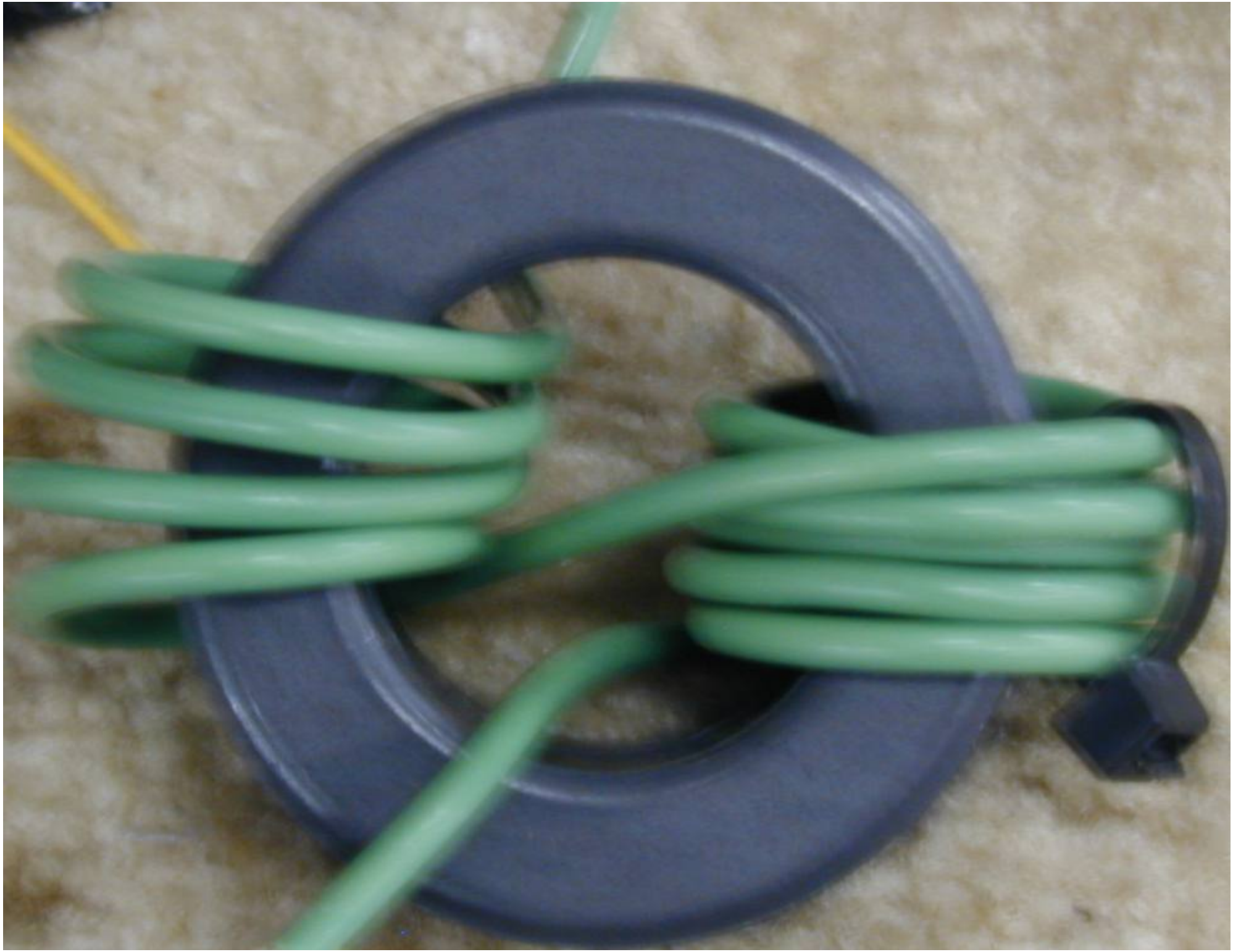
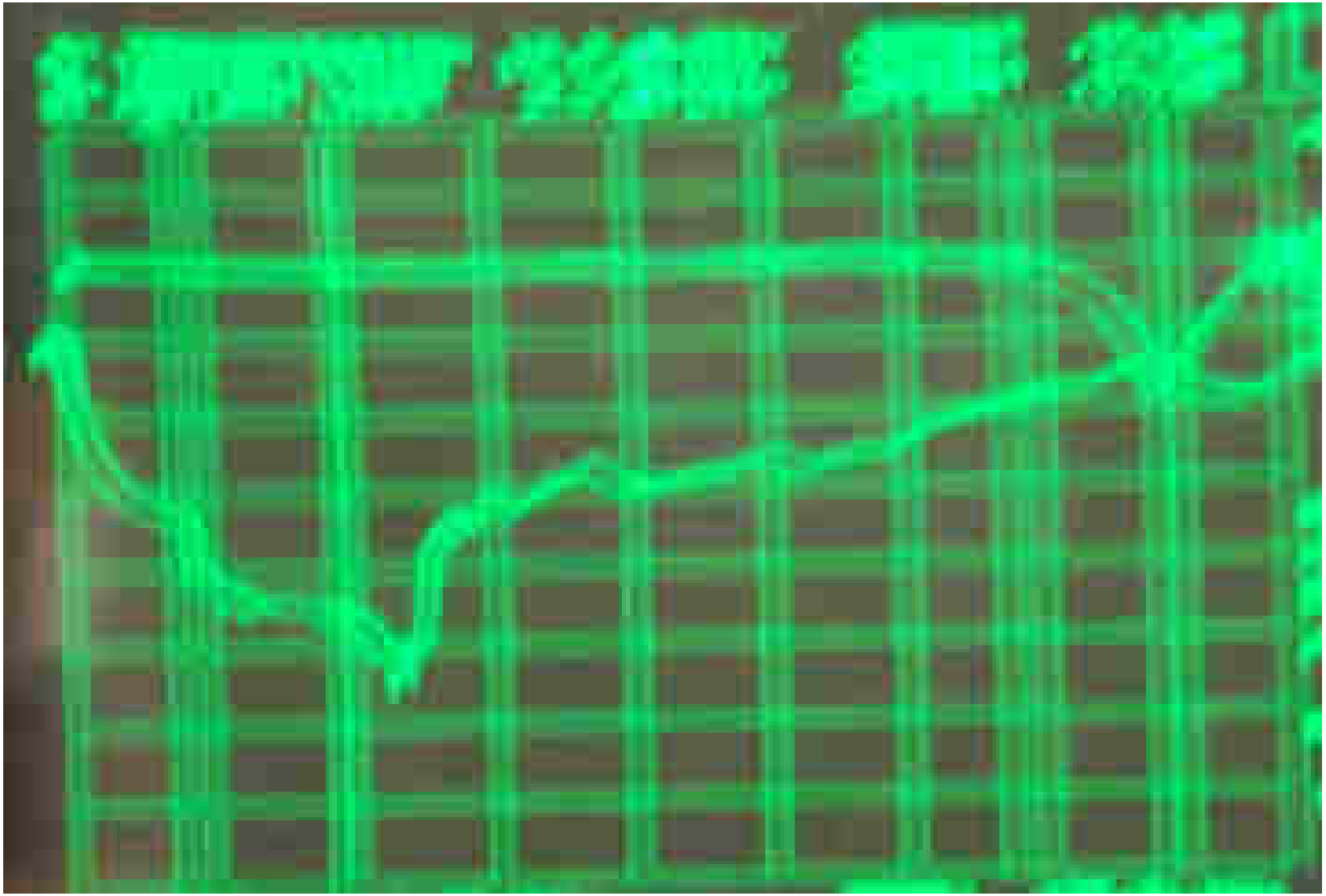
Toroid Chokes and Baluns

Commonly published information tells us winding a choke or 1:1 balun on a toroid with this special winding technique increases common mode impedance (choking isolation) of an 11-turn balun on a pair of #65 material ferrite cores:



The common mode rejection sweep, assuming a 50-ohm common mode source and load is:

(pardon the camera jitter)

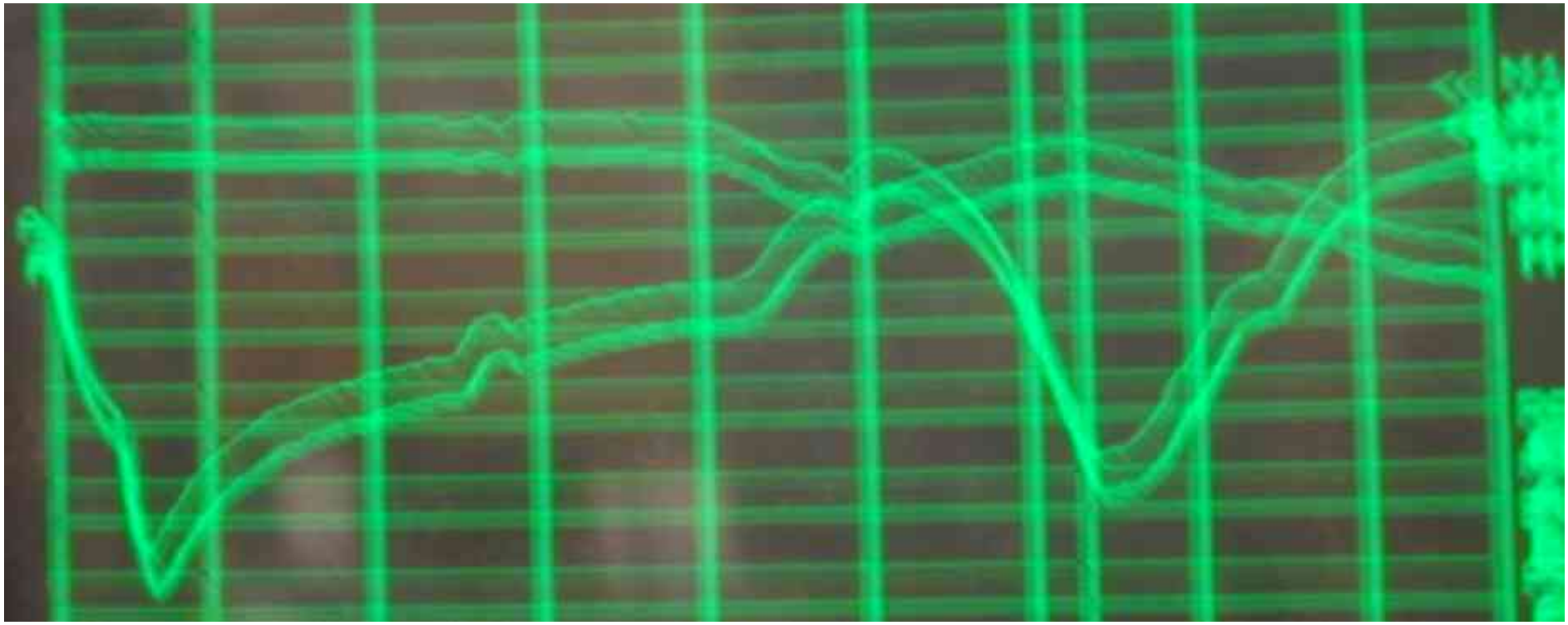


This sweep shows the following suppression using split winding method:

15.71 MHz	37.1MHz	134.8MHz	157 MHz	
-33dB	-46dB (peak performance)	-19dB	-3dB (minimum rejection)	

The same core was rewound using a normal parallel winding:

The sweep looked like this:



Common mode rejection, compared to the more complex winding, was:

Frequency	15.71 MHz	37.1MHz	134.8MHz	157 MHz
Split winding	-33dB	-46dB (peak performance)	-19dB	-3dB (minimum performance)
Normal winding	-47dB (peak performance)	-41dB	-4.6dB (minimum rejection)	-11dB

Conclusion

The ONLY change by using the more-difficult split winding is peak performance points moved higher in frequency. Performance in the HF range and lower actually decreased, so the extra effort resulting in a poorer HF balun.

This follows what many of us have been saying, and what anyone familiar with resonant circuits already understands intuitively. Common mode impedance increases with a slight amount of parallel capacitance is adding. The capacitance does NOT, as some claim, reduce performance by "leaking" RF around the choke.

The rule is simple. With any parallel L and C (neglecting losses), impedance always increases unless the parallel capacitive reactance (in ohms) is less than half the value of the inductance (in ohms).

With a choke winding impedance of 300j with an added parallel -150j capacitive reactance, the combined impedance would be -300j. The isolation would be the same! Any value of capacitance reactance higher than that (less capacitance value) would actually increase choke isolation!

There are many "improvements" we just blindly accept from reading articles, because on the surface they seem to make sense. This dispels one of the myths about baluns.

Another common myth is the widely accepted claim moving a choke balun to a lower common-mode impedance point of a circuit, such as the input of a tuner, improves balance or balun performance in the system. If you think baluns at the input of tuners are a good idea, go to this W7EL link and read the section on current baluns. You might be surprised!

Inductors and Loading Coil Current (Mobile and Loaded Antennas)

[Home] [Up] [Independent Measurements] [Inductor operation]

Related pages:

Mobile antenna FS comparisons Loading Inductors Inductor Spice Model

W7EL's Measurements and comments Constructing an RF Current Meter

Much of the data below also applies to inductors in equipment, such as tank circuit.

If you arrived here from a link from K3BU, *welcome to W8JI.com I hope you enjoy your visit.* ***If you are not linked here from loading coil current articles on another web site, please skip the blue-text immediately below. It is meaningless.***

If you arrived here from a link from K3BU, I want to apologize for the personal nature of that page and this response. Why is this response necessary? None of us like to see false or out of context claims or personal attack articles. Many of the comments from below were removed from context and quoted to mean something else in another internet article. It is important to read everything ***IN CONTEXT****.*

How Does an Inductor or Loading Coil Work?

The most basic answers are:

What does the coil do? A loading coil does not replace a missing fraction of a wavelength. The coil simply inserts a series inductive reactance that cancels capacitive antenna reactance. When a 150 ohm reactance inductor is inserted in series with a 150 ohm capacitive load (like an antenna), only the resistive parts remain.

What determines current distribution in a loading coil? The capacitance to the outside world and the impedance above the loading coil. The current in any inductor would be equal at each end except for displacement currents, which are "imaginary currents" that flow through capacitance.

How much difference is there in loading coil current entering the coil and loading coil current exiting the far end? If the antenna beyond the coil has a low self-impedance compared to the impedance of the shunting capacitance from the coil to "ground", the currents at each end of the coil will be essentially equal. It has NOTHING to do with electrical degrees the coil compensates for. In other words if the portion of antenna above or beyond the loading coil is long or has a large area compared to the physical size of the coil, current is essentially equal throughout the coil.

What does significant current taper in the loading coil indicate? It generally indicates a poor antenna design, where the loading coil (or any other application using an inductor) has high stray capacitance to other areas of the antenna system (like the groundplane) compared to the capacitive reactance of the antenna beyond the inductor. Significant current taper indicates a poor loading coil or poor antenna design.

click to view typical installation measurements

The Difference Between a Loading Coil and a Normal Inductor

There really isn't any difference, except the location where the coil is used. Mobile loading coils or loading coils for short antennas often have very high reactance. They have small amounts of capacitance at the end, and so stray capacitances are more of a concern. Stray capacitance from turn-to-

turn increases circulating currents and has the effect of increasing inductance and effective resistance at the expense of reducing bandwidth and Q. Stray capacitance to the outside world causes the coil to behave like an L-network, and transform impedances instead of providing a series reactance. This is why the optimum form factor of a coil becomes longer compared to the diameter with any inductor having very high reactance.

Inductors with low reactance are less critical of stray capacitances. Optimum form-factor in a low reactance inductor leans towards a short coil with the diameter nearly equaling length. In tank circuits or loading coils used with longer antennas or with capacitance hats, an optimum inductor is shorter and larger in diameter.

Most optimized inductors fall between 1:1 and 4:1 length to diameter ratios, the exact value depending on the terminating reactances. A coil is a coil, they behave the same way regardless if used as a loading coil or a tank inductor.

Common myths about inductor behavior:

One common myth is loading coil current is reduced as it passes through the coil. There are two reasons cited for this. One idea is the current is reduced because the loading coil replaces a certain amount of "electrical degrees" of antenna area, like the current taper in an antenna. The other idea is that series loss resistance causes a current reduction.

We often find inexperienced builders of 5/8th wl antennas think the "loading coil" needs to contain 1/8th wavelength of wire in order to make a "3/4wl antenna" and thus cause a low feed impedance. In other cases, some claim a half-wave of wire wound on a form causes a 180-degree delay, and is useful for phasing in a collinear array.

The basic flaw is the above ideas do not account for what happens in a coil with mutual coupling between turns. The flawed viewpoint is that current goes in one end, winds its way around through the physical length of wire in the coil, and after a time delay comes out the other end. There is a physical mechanism that prevents this, as we might intuitively think, from actually happening . The mechanism is the magnetic field in the coil!

What Really Happens

When current flows in the transmitter-end of the coil, a magnetic field is created. This magnetic field causes charges in the other turns to instantly move. This effect ripples through the coil at light-speed, just over 186,000 miles per second. As long as the magnetic flux coupling is high, the delay through the coil is the speed of light over the physical length of the coil. The electrical delay is the physical length of expressed the coil expressed in degrees at the operating frequency.

(Another interesting effect occurs. The increasing magnetic field sets up an "opposing voltage" as it cuts across conductors. This opposing voltage, created as the field expands, is what causes the current to rise slower than the applied voltage. If the exciting voltage is decreased the field collapses, and now the voltage changes polarity and aids current flow! If we don't allow the current to flow, the voltage will rise until it does. This is what causes the kick in a relay coil when we open the relay coil path, or the spark in an ignition when the points abruptly open.)

In an RF system, the physical size of the coil actually does add some "antenna effect". For example, on 160 meters the wavelength is about 550 feet. 1.5 feet is about one electrical degree. A skinny one foot tall coil, with negligible stray capacitance, would have about 0.67 electrical degrees phase delay. This delay occurs because to coil occupies a physical length of .67 degrees. Current at each end would be almost perfectly equal, the taper would be about what we would expect for a fractional-degree-long coil.

(*In the real world, all components have some stray capacitance and flux leakage, so they have a different amount of electrical length and current taper than the "negligible capacitance" case. In good coil designs, the capacitance and leakage is small and can be ignored. I'll show you measurements later to prove this.)*

Now let's look at an extreme case. If the entire antenna is "coil", like a helically wound antenna with no top hat or stinger, current would be reduced to nearly zero at the open end. This is because distributed capacitance over the length of the antenna is fairly high, the shunting capacitance has a low impedance compared to impedance at the end of the antenna, and current is diverted to ground in the form of displacement currents.

Compact loading coils are another matter. In many cases phase delay is negligible or immeasurable by normal methods...flux coupling is nearly perfect. A good example would be a relatively compact toroid or a compact nearly-square L/D ratio loading inductor. I've found it impossible to measure the current taper in a toroid and very difficult to measure in a compact air-core loading coil. (The opposite extreme would be be a perfectly straight wire with no folds or bends or the helical antenna described above.)

In the case of the toroid or compact coil, the behavior would be such that doubling the turns nearly perfectly quadruples inductance. If we doubled turns and inductance simply doubled or increased at a much faster rate, we should know the coil is in a mode other than a pure inductor mode. This is a ***strong*** indicator inductor operating Q is less than optimum, and the inductor might behave less than ideally in critical applications.

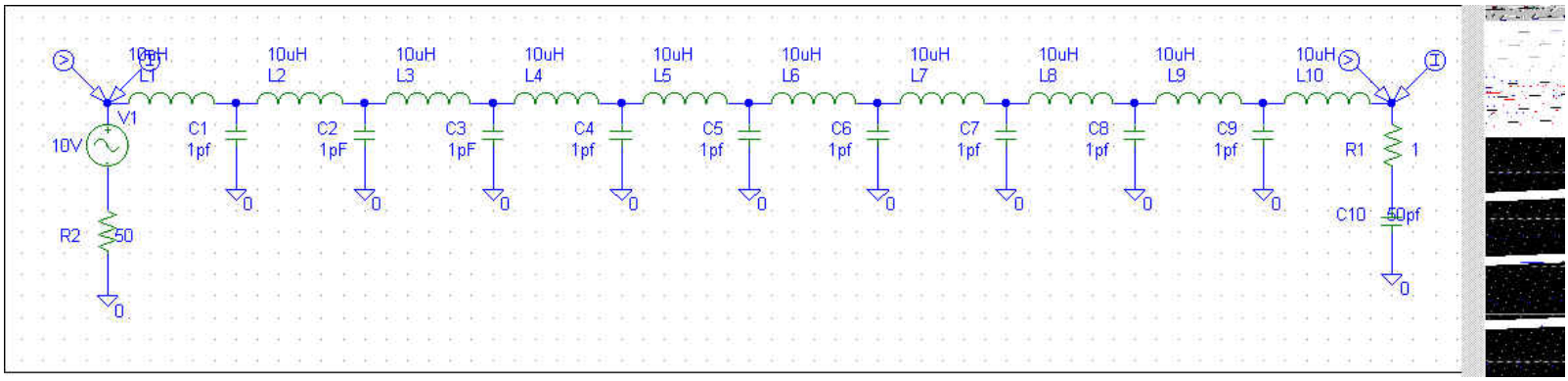
As a matter of fact, observing inductance change while adding turns can be an excellent test for flaws or shortfalls in system design. A linear increase in inductance when adding conductor length indicates design problems.

A perfect impedance squaring effect indicates minimal electrical phase-delay, or "antenna length" of an inductor. Impedance squaring as turns are doubled indicates the undesired inductor stray capacitance has a high reactance compared to the antenna system beyond the loading coil. Of course there can be exceptions, but it is a good general rule that large current taper indicates the loading system is much less efficient than necessary.

Making a Delay Line

It's certainly possible to make a delay line from a coil without opposing flux, but doing so requires stray capacitive reactance to be significant compared to the value of distributed inductance in the coil. This would occur in a very long helice, a very large diameter helice or loop, or an inductor near a large metal counterpoise or ground plane.

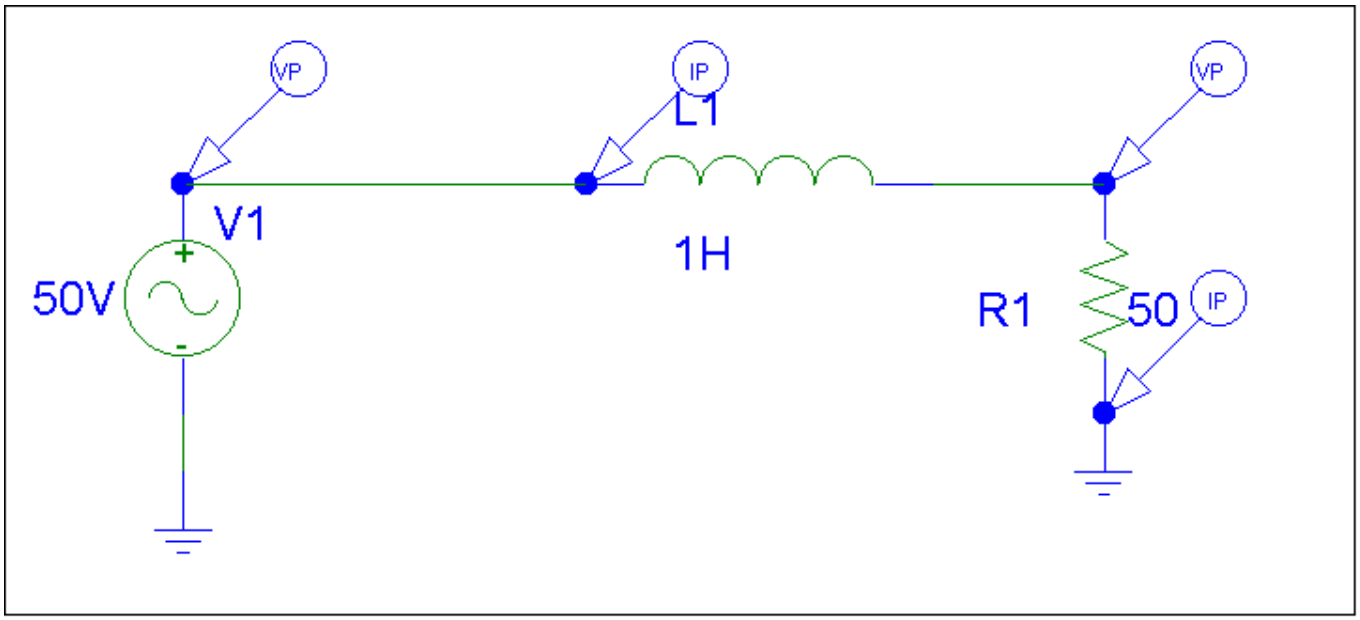
It's important to remember unless a coil is "stretched out" or "expanded" a great deal, the phase delay will not even be close to the physical conductor length. (The exception could be if you had so much capacitance the inductor acted like a series connected string of L/C/L networks as shown below).



In any case while this effect might be good in a collinear antenna or plate choke (assuming you do it right) it is a BAD effect in a short loaded antenna!

Inductor E/I Phase shift

An inductor delays the flow of current in relationship to applied voltages as the magnetic field inside the coil expands. Voltage increases before current starts to flow. This phase relationship between voltage and current is often confused with time-delay phase in the inductor. Say we have this simple circuit:



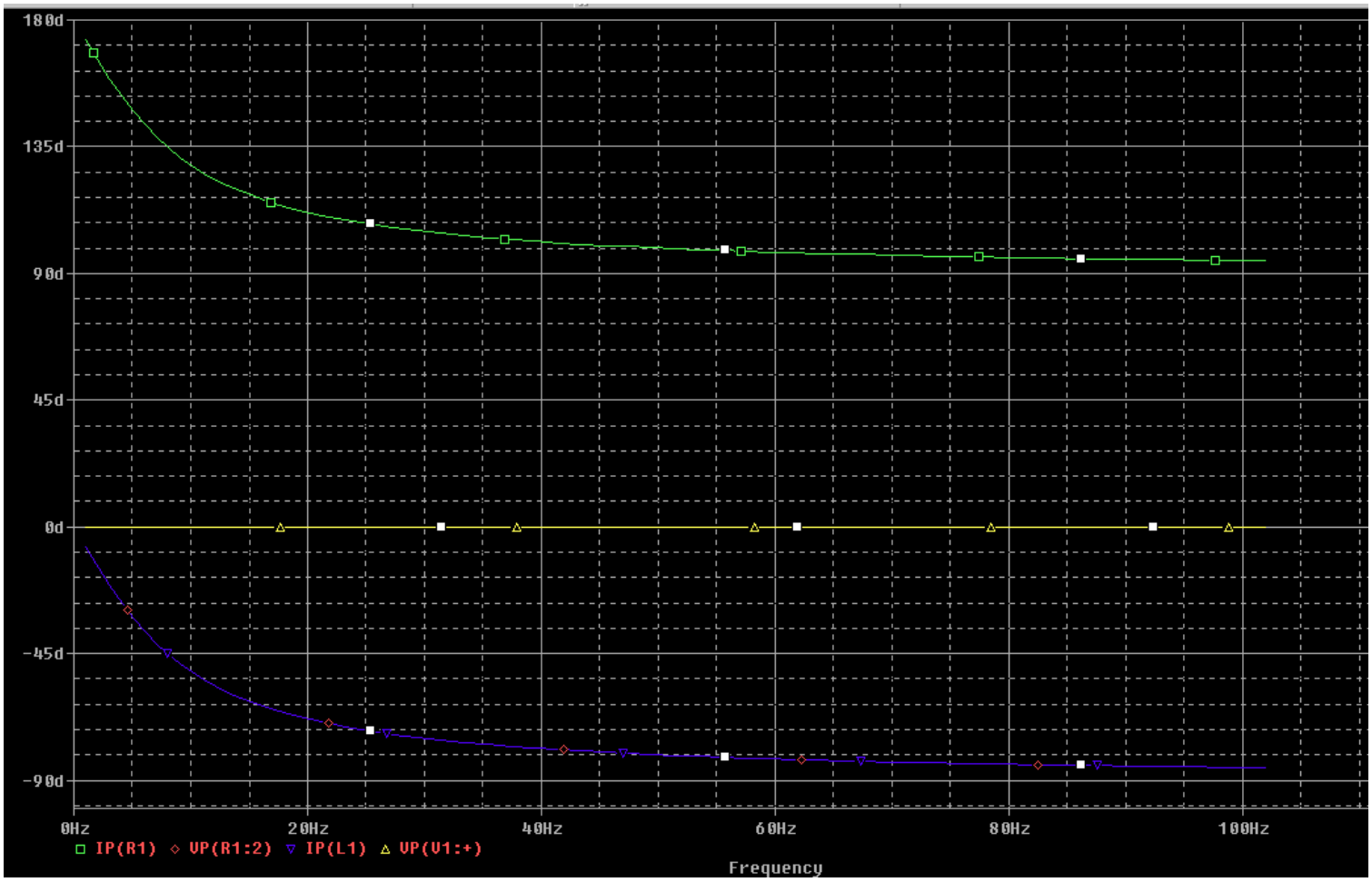
Current and voltage at V1 will be out-of-phase by the effect of L1 "charging" with magnetic flux. Current appears AFTER voltage rises, and falls after voltage falls. Current in R1, however, is exactly in phase with voltage across R1. That's because the voltage across R1 is always E=I*R.

Every component must follow the Laws or electrical rules established for that component.

The current in R1 is delayed from VOLTAGE rise in V1 by the voltage to current phase delay of L1. This does result in a time delay in relationship to voltage rise at V1, but there is NO current time delay through L1! V1, L1, and R1 all have the same peak current at the same time!!!

The notion that antenna loading coil delay current by the same time as they delay response to increased voltage is obviously nonsense.

Here is a graph of phase delays in the above system:



Current in the load, generator, and inductor all exactly track in the same relationship from dc up. There is no "phase delay". The generator voltage is a straight line different than current, and this indicates the generator sees a "reactive load".

The Misplaced Notion

Proponents of the idea that coils replace "antenna length" so far have been unable to define a set of rules or logical reasons why a current reduction and electrical-degree phase delay related to antenna area "replaced" would occur in a two-terminal component. While a long inductor with poor flux linkage from end-to-end or an inductor with low values of stray capacitive reactance to a groundplane compared to series impedance can cause SOME current inequalities or phase delays, the amount is normally immeasurable with normal thermal current meters with a reasonable coil form factor and termination above the coil. The amount of current taper actually rivals the disturbance of the system by adding the measurement device, unless we are very careful in how we construct the measurement device.

Both W7EL and myself made independent measurements that show the fact the coil or inductor is in an antenna has nothing to do with current distribution in the coil. Some people have actually incorrectly reported W7EL's data! Here's what he had to post on rec.radio.amateur.antenna to correct misrepresented claims.

The Need for a Measurement

An article on E-ham claimed measurements proved a new concept about loading coil current. The E-ham article put forth an idea that current disappears as it moves through a loading inductor without a mechanism like displacement currents providing a path. This claim conflicts with established component behavior, so it would indeed be fascinating if it were true! One of the claims supporting the idea that coils in antenna work differently than coils in circuits was that a non-radiating toroid loading inductor showed a current taper when used in an antenna.

I recently constructed a calibrated current meter that slips over whip antennas and masts, and is for all practical purposes totally immune to variations in voltage in the system. It also is mostly plastic, and has minimal effect on stray capacitance of the antenna. The resonant frequency and currents are not significantly perturbed by measurements with this meter. When I added a similar meter used in the other tests, resonant frequency shifted significantly! This is a sure sign the meter's capacitance or inductance is affecting the system.

In late December 2003 and early January 2004, I made additional measurements of loading inductor currents. The results clearly agree with the analysis that had been presented on this page since early 2003. Without displacement currents, currents into and out of a loading inductor *are equal.* That is a hard rule, it agrees with theories defined by people much smarter than me, and I believe it is unbendable unless the works of Faraday, Maxwell, Ohm, and Kirchoff were incorrect.

A sample of measurements above and below the loading coil with various antenna above the coil (current as percent of reference) follow:

	Toroid with hat	Small 2x2" coil with 24" hat up 24"	Long 12x3" coil with 24" hat up 24"	Long 12x3"coil with 6' whip
Current below	100%	100	100	100
Current above	100	94.4	73	76%

	Toroid with whip	long 12x1.5" coil with 6' whip	long 12x1.5" coil with 24" hat	Small 2x2" coil with 6' whip
Current below	100%	100	100	100
Current above coil	100%	79%	75%	96%
Current in whip 1ft above top of coil		73%		

The most revealing thing was how noticeably small changes in stray capacitance near the middle and top of the loading coil affect current distribution. It was quite evident hanging a large meter on each end of the coil would greatly perturb the system.

Clearly we do NOT want:

- **A large hat just above a large coil**
- **A long large coil and a short whip**
- **A coil near large sheetmetal**

More data along with photographs will appear on a new page over the next month or so. Until then, I can assure everyone the conventional theories presented below are accurate, and the theory that "electrical length" the coil "replaces" is incorrect. Loading coils indeed behave like any other inductor in the world.

Independent measurements by a reliable engineer have agreed with my measurements above. Anyone doubting my data need only read the following e-mail from W7EL.

On to the old text in this page:

The Incorrect Assumption

Another commonly misconception is, since voltage increases at the far end of the loading coil, current must logically decrease. After all, we have a fixed amount of power and voltage has increased. The assumption is:

1.) We multiply voltage times current to get power.

2.) If voltage increases current must decrease.

Unfortunately, this is ***not*** correct in reactive systems! Simple P (power) = I (current) times E (voltage) only works when the system is non-reactive. This condition only occurs at resonance, and only ***below*** the loading coil at the antenna feedpoint!

In a reactive system, like in a mobile whip above a loading coil, voltage and current are no longer in phase. As a matter of fact, voltage and current can closely approach being 90 degrees out-of-phase when the whip is electrically very short. Since the antenna area above the loading coil is highly reactive (voltage is not in phase with current), we can not multiply voltage times current without considering phase differences.

You may have heard the term "reactive power" or VAR (volt-amperes-reactive). Reactive power is voltage times current without consideration of phase angle. We can have kilowatts of VAR power with only a low power transmitter, and that is what we actually have in the reactive part of the small antenna.

Coil Q and Changes in Efficiency

Current taper or reduction has been cited as a reason coil "Q" has little effect on signal level in mobile systems. Speculation is only the first few turns of the loading coil carry significant current because the coil "leaks" magnetic fields and radiates, and this is why the coil Q has little effect.

Another idea proposes the loading coil "makes up" a certain missing part of the antenna. It goes on to conclude the loading coil can be accounted for in "electrical degrees", making up the "missing difference" in antenna degrees. This isn't true either. The inductor doesn't know where it is and suddenly change from "x" ohms reactance to electrical degrees! It responds to AC currents and voltages as any inductor in any circuit does. It doesn't suddenly change measurement units.

As an example of this, try to define a 45-degree electrical length inductor at 1.8 MHz. That would mean it is a capacitor at 3.61MHz, where it is over 90-degrees long! How many turns at what length and diameter is a 45-degree inductor?? Where is a formula that allows converting a given size inductor to electrical degrees? This shows how useless and meaningless that definition is!

The inductor adds a certain amount of series reactance, that's all. A 300uH inductor is not 20-degrees long, nor is it 80-degrees long, so far as radiation goes unless it is really that long ***physically***. It is a certain number of ohms reactance at a certain frequency, or a certain number of units called Henries. It is not "electrical degrees" that it adds, it is a non-dissipative reactance (in combination with a loss resistance because of finite quality) at a certain frequency!

A loading inductor can "insert" a large amount of phase shift, but the phase shift is between voltage and current. The only exception to this would be if the inductor had considerable distributed shunt capacitance to the outside world, and acted like a string of series inductors (with the antenna) and shunt capacitors (shunting to the ground system). In that case we could expect coil Q to be extremely low, since it would be the electrical equivalent of a lossy transmission line. That's either an awful loading coil, or it is a less-efficient helical loaded antenna!

The Correct View

Another group of people don't argue against established and proven circuit theory. They understand charges flowing into one end of the loading coil must have someplace to branch off (a virtual third terminal), or they must flow out the other end. Without that additional "virtual" path, charges flowing into the coil would always equal charges flowing out. This is true regardless of radiation, losses, or induction fields.

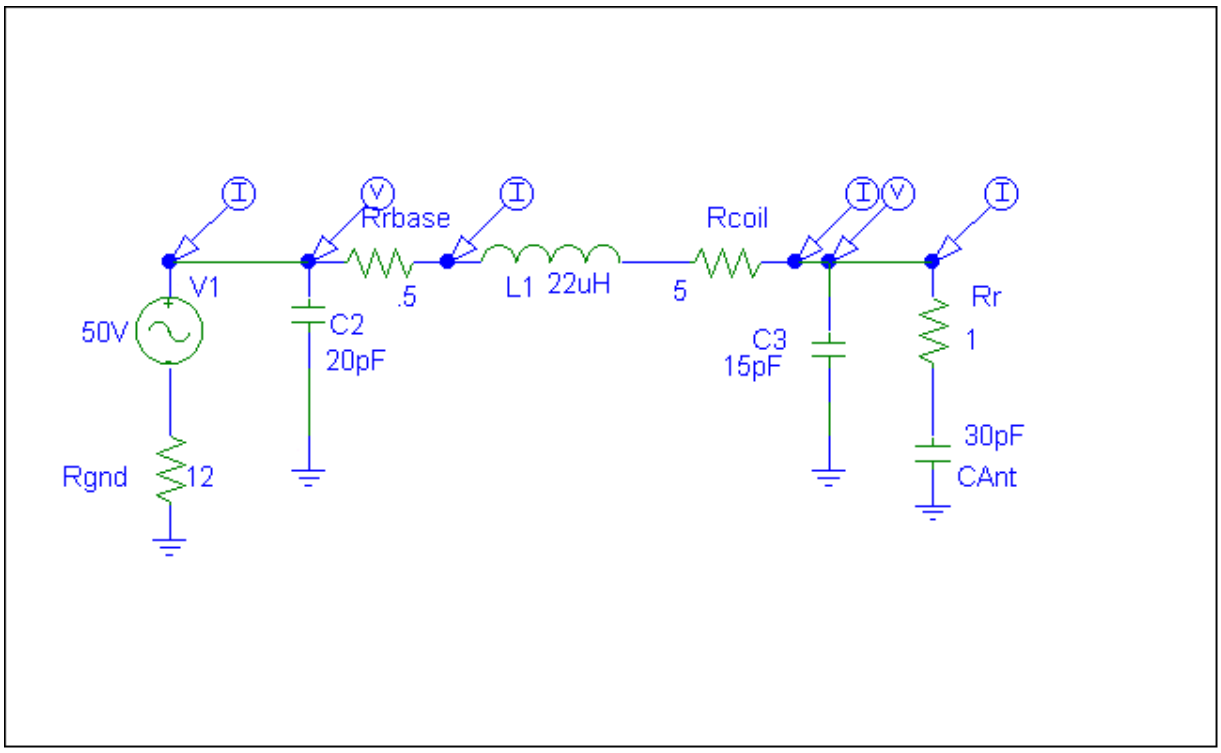
This makes perfect sense when we think of any dc circuit, antenna, or RF system. Electrical rules are satisfied, the system behaves as it does in the real world.

There is very little change in current, unless the coil is physically very long compared to the rest of the antenna above the coil or unless the coil is laid right against "grounded" conductors and the whip above the coil is very short. This fits perfectly with helical verticals, where the coil is "stretched out" over the length of the antenna.

It also agrees with base loaded antennas, which have nearly as much current into the antenna above the coil as at the feedpoint. It agrees with center loaded antennas, where current below the coil is essentially uniform and the whip above has triangular distribution.

Current ***can*** be different in various areas of an inductor, but ***only*** if shunting capacitances (impedances) to the outside world are significant compared to load capacitance (impedance). Another condition where current can vary substantially is with operation near the condition of self-resonance in what is normally considered or defined as a "series-resonant" mode. This would be a very poor and inefficient loading inductor, such as when a 160-meter antenna is used at a secondary resonant frequency in upper HF.

Circuit Model of a Mobile Antenna



The above model shows what might be a typical mobile antenna installation.

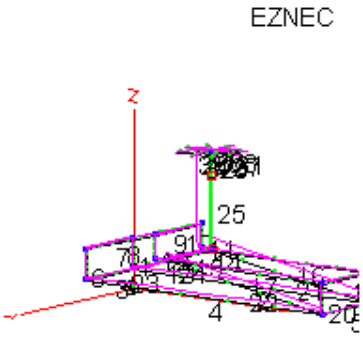
1. Rgnd ground resistance of vehicle normalized to feedpoint
2. V1 coaxial feedline
3. C2 base capacitance
4. Rrbase Radiation resistance of the base area of the antenna
5. L1 loading coil
6. Rcoil coil equivalent series loss resistance
7. C3 coil shunt C to ground
8. Rr top area radiation resistance
9. Cant equivalent antenna capacitance above coil area

My 160 Mobile Antenna

I've worked all continents except Africa while 160-meter mobile. I have CW contacts at over 10,000 miles, and SSB as far as Europe (4000 miles). My mobile antenna consists of an eight- foot antenna with a six-foot hat (made from surplus Ford or Chrysler car antennas) at the top. The loading coil is at 5 feet.

This antenna has been on the truck for thousands of miles, without mechanical failure. It is mounted at the upper left corner of the truck bed, about one foot back from the cab.

The following is a model of the current antenna system on my Ford F-250 HD long bed super cab truck:



The base impedance in the model is:

Frequency = 1.854 MHz.
Source 1 Voltage = 24 V. at 2.35 deg.
Current = 1 A. at 0.0 deg.
Impedance = 23.98 + J 0.9853 ohms
Power = 23.98 watts
SWR (50 ohm system) = 2.087

Actual measurement at my Johnstonville, GA farm in open flat pastures on August 17 at 8AM. Wet soil 25.8 ohms 0j base impedance, pretty close agreement to EZNEC model and earlier data! (I did have to adjust the model for very low ground conductivity, otherwise the resistance was far too low. It's my belief that NEC-2 underestimates ground losses in small radial or counterpoise systems that are close to earth.) Earlier text shows a base resistance of ***28 ohms***, that was dry soil with a slightly different loading coil and antenna.

The modeled current distribution for 1-ampere applied at the base (in 1-foot intervals) is:

1ft= 1.0031

2 ft= 1.0091

3ft= 1.0178

4ft= 1.0318

<Coil>

5ft= 1.0175

6ft= .97512

7ft= .92984

8ft = .89522

Measuring the current into and out of the loading coil with a small thermocouple RF meter, I detect no difference This is in close agreement with the model.

The efficiency of this antenna knowing coil Q, radiation resistance, and base resistance calculates just under 1 percent. The model indicates about 1/3 percent efficiency. This is reasonably close.

Removing the hat (in the model only) shows the following changes:

1ft = 1.0043
2ft = 1.0133
3 ft= 1.0279
4 ft= 1.0566

<coil>

5 ft= .95508
6 ft= .72232
7 ft= .27813
8 ft = open

I haven't tested the above, but with the same loading coil loss resistance the model says efficiency is now around 3dB worse. Removing the hat, with NO change in coil resistance, shows nearly loss nearly doubles. Of course the coil resistance would increase, because the loading coil nearly quadruples in size. Bandwidth is less and efficiency is less, even if I could maintain the same coil resistance.

Examples of Unequal Current

In the above models, we see that current into and out of the one-foot long coil is about the same. There is only about ***2% change in current*** even though the coil occupies 12% of antenna length in the "hat-loaded" antenna, but in fairness I couldn't resolve that change with a reasonably good RF current meter.

The model predicts ***10% change in a non-hat antenna***, but I never measured that antenna to confirm it.

Clearly there is no basis to the claim current is high only in the first few turns of an inductor, or that current tapers in relationship to "electrical degrees". The most accurate way to state the effect would be to say: "When the loading coil is short and the capacitance of the antenna beyond the coil is reasonable (in this case 3000 ohms Xc or less), there is an immeasurable reduction in current in the coil. When the required loading reactance is very high (in this case 8000 ohms), the reduction in current is about what we would expect for an equivalent length of antenna replacing the coil."

Degrees Vs Radiation Resistance

This upper four feet of this antenna resonates near 24 MHz with the hat. We can assume it is 90 degrees long at 24 MHz, which would translate to 6.9 degrees on 1.85 MHz. Following that same logic, this would mean the loading coil would be about 83 degrees long electrically. Using the incorrect logic proposed by others where the loading coil "makes up the difference in electrical degrees", there would be almost no current past the loading coil. Obviously this is not the case, the loading coil has very little "electrical length". As a matter of fact, the electrical length is about equivalent to the physical length!

This goes back to radiation theory, and my favorite saying: "Five hundred feet of wire in a one foot long tube is still one foot of antenna". Some CB manufactures sell antennas to consumers with the claim they use 5/8 or 3/4 wavelength of wire in an eight-foot fiberglass whip, so the antenna has more gain. Obviously this is not true. Let's not let such silly claims spread into amateur radio!

Related topics:

Inductors

The spice inductor model shows one example of how unequal current is created. The model demonstrates a coil having significant distributed capacitance to the point of current return in the system compared to terminating impedance of the coil. In a monopole this return path would be to the groundplane, or anything closer to the potential of the groundplane than the area above the loading coil's position in the antenna system.

Another Practical Antenna Example

Let's assume we have a lossless 15.3 foot long 0.2 inch diameter conductor over a perfect groundplane. Eznec gives the 1.821 MHz base impedance as .3004 -2169j. In other words, the antenna "looks like" .3004 ohms of load resistance in series with 40.32pF on 1821kHz. The return path for current is through the .3004 ohm resistance and 40.32pF capacitance, back to the ground of the antenna (it is a Marconi antenna).

Such a termination (load) would require a series inductance of 2169j (189.57µH) to cancel feedpoint capacitive reactance. A typical 190µH inductor

would be rather large, requiring somewhere around 53 turns when using a 4" by 4" form factor. One would expect a physically large inductor to have noticeable but very small displacement currents to the groundplane, when the small stray coil capacitance is compared to the 40.32pF termination capacitance. This raises two very important design guidelines:

- ***When installing a loading coil of substantial inductance in an electrically short antenna, sheetmetal and dielectrics should be kept away from the coil and areas of antenna above the loading coil. This would include dielectrics on or near the inductor, since the presence of dielectrics would increase undesirable capacitance.***
- ***When inductive reactance requirements are large, as when short thin "stingers" without hats are used above a coil, the coil form factor should lean more towards long and thin. Capacitances near the open end of the coil (high voltage end) should be minimized. This would be true even when the coil length increase results in a small reduction in mutual turns coupling, since the stray capacitance may result in a larger loss penalty than the slight increase in accumulated resistance from additional wire length.***

Efficiency

Efficiency in any antenna near earth is almost always dominated by ground related losses, short-height Marconi antennas are no exception. The overall effect of loading inductor Q and matching system losses are "diluted" or "swamped-out" by ground losses. Ground losses cause most systems to have greatly reduced sensitivity to inductor design.

The only consistently predictable factor in efficiency in fractional wavelength Marconi antennas with limited size ground systems is radiation resistance. Efficiency increases almost directly in proportion to radiation resistance.

Radiation Resistance and Power Radiated

Radiation resistance is probably the most poorly defined term used with antennas. The lack of clear definition creates errors and misjudgments when predicting antenna performance. If you wish more detailed information, this page contains information on radiation resistance. For the purposes of this discussion and to avoid pitfalls associated with using feedpoint impedance as radiation resistance, I'll use the same definitions Jasik, Balmain, and others have used. This definition is based on the IRE definition of radiation resistance being equal to the net or effective current causing radiation squared divided by the power radiated as EM energy, or Rr=Pr/I^2.

Using this definition, a folded dipole has a radiation resistance identical to a conventional dipole of the same physical dimensions (~70 ohms).

Radiation is caused by charge acceleration, there is no magic. The only thing affecting radiation resistance in a short vertical antenna near ground is current distribution over the linear area occupied by the radiation portion of the antenna. The general rules are:

Radiation resistance of a Marconi vertical in the maximum possible radiation resistance case for a given height (this is the case where current is uniform throughout the structure) is equal to 1580*(H/L)^2 where H equals height and L equals wavelength and both are expressed in the same units. Using degrees, we see a 10-degree tall antenna has a maximum possible radiation resistance of 1580*(10/360)^2 or 1580*.000772 = 1.22 ohms. This would apply even if the antenna is a vertical, DDRR, Fractal, or folded unipole with considerable top loading.

If current is triangular, radiation resistance would decrease by a factor of four to 0.305 ohms.

Power radiated is given by I^2*Rr

With 100-watts applied to a 10-degree tall antenna, net current in a lossless antenna with uniform current distribution would be 9.05 amperes. With triangular distribution, such as appears in a small diameter short base loaded whip, current would be approximately 18.1 amperes. We are in serious problems if the inductor reduces current along its length, since the only possible way to radiate 100 watts would be to have somewhere around 9 amperes of effective current integrated over the 10-degree vertical area of space for the radiator!

Ground Losses

All current flowing (or displaced) vertically into the antenna must equal current flowing out of the ground or counterpoise system. Even though ground losses are distributed losses, we must normalize all losses to the feedpoint in order to compare systems. There are cases where this will not always occur, causing us to falsely assume we have lower losses than really exist.

In this tutorial and comparison, I have normalized ground losses to the same point where radiation resistance is considered.

System Losses

(Measured data below of actual antenna given below was from 1995 data taken at a different location near Atlanta with a slightly different loading coil and antenna. There is a slight disagreement with current data. I left this all in so you can see the departure from measurements and models using 8 year old data.)

Base Loaded (Triangular Antenna Current Distribution) with no ground loss

Assuming we have a base-loaded antenna, and the operating frequency has a wavelength of 550 feet (around the 160-meter band), a 15.3 foot vertical would fit the above 10-degree value. Interestingly enough when we compare Eznec to formulas available in older (1950 vintage) engineering textbooks, we find radiation resistance predicted by Eznec is .3003 ohms while the triangular current estimate for the same height radiator is .305 ohms! This is an amazing degree of agreement, illustrating what we could do before modeling programs became available. (With perfect top loading, both Eznec and longhand calculations show approximately 1.2 ohms of radiation resistance.)

Assuming our 15.3 foot tall (10-degree) base-loaded antenna uses a coil Q of 200, the coil has 10.845 ohms of ESR. Total resistance with a perfect ground would be 10.85+.3= 11.15 ohms. Current into this system with 100 watts applied would be around 3 amperes, resulting in ~2.7 watts radiated and ~97.3 watts lost as heat in the inductor.

Doubling coil Q (400) would provide 5.73 ohms of base resistance with 4.18 amperes. Power radiated would be 5.2 watts, power lost as heat would be 94.8 watts. Efficiency does not quite double, changing from 2.7 to 5.2%. This results in a 2.8dB change in signal level.

Top Loaded (with no ground loss)

If we added a four-wire hat with 15-foot wires, current would no longer be triangular. While we wouldn't quite reach the optimum uniform distribution, current at the top would be about 78% of current at the antenna base. Feedpoint impedance would become 0.97 -551j, and the antenna would look like 0.97 ohms in series with 159pF.

Using a coil Q of 200, we would now have 2.76 ohms of inductor loss. Current becomes 5.18 amperes. Radiated power is 26 watts, while power lost as heat becomes 74 watts. Even in the perfect ground case, the change in efficiency caused by top loading is large. Top loading (with only the hat) results in 9.8 dB change in signal level when compared to the base loaded case when coil Q remains 200. Efficiency is 26%. The coil remains at ground level for easy matching and frequency change.

In this case current at each terminal of the loading coil would be essentially the same regardless of poor coil mounting techniques. In order to have significant current taper in the coil or in the bottom of the mast, shunt capacitance would have to be a significant compared to 160pF. The antenna's high input capacitance relaxes inductor and antenna mounting electrical requirements.

Base Loaded (high ground loss)

My F-250HD Super Cab pickup truck, when parked over open medium quality pasture land, has a ground resistance of about 20 ohms (normalized to the feedpoint) on 160 meters. Applying this ground loss to the base loaded antenna, the system has a feedpoint resistance of 20+.3=20.3 ohms. (This is reasonably close to actual feedpoint resistances measured with a similar operating antenna.) Adding coil losses, the system has 20.3+10.85=31.15 ohms. (**NOTE: Current coil is ~8 ohms ESR, 10.85 ohms is from ~8 year old data**) Current is sqrt (100/31.15) or 1.79 amperes.

This results in .96 watts radiated, and 99.04 watts lost as heat. Efficiency is now around .96%.

Substitution of a coil with a Q of 400 results in 25.7 ohms feed resistance, or 1.97 amperes antenna current at 100 watts. In this case efficiency is now 1.16% for 1.16 watts radiated. The change caused by doubling coil Q with high system ground losses is about 0.8dB, compared to almost 3dB in the perfect ground case! With a poor ground (in this case typical of a very large vehicle), a large change in coil Q produces little change in system efficiency.

Another Top Loaded (high ground loss) System Example (made prior to the EZNEC model above)

Using a large hat isn't practical in a moving mobile, although it could apply to fixed stations suffering with poor ground systems. When the hat is smaller, such as a mobile requires, the loading inductor can be moved higher in the system. Such a move would produce uniform current below the loading coil, with a current shape above the coil dictated by the construction of the upper portion of the antenna. My own mobile uses a six-foot diameter hat manufactured from stainless steel automobile antennas arranged in a spoke. I have no problems with wind or occasional obstructions. While unsightly, a modest hat is workable.

In order to keep the systems comparable I'll use the same radiation resistance provided by a large hat, but intentionally add high ground loss as a lumped resistance. This model ignores field losses near the antenna.

In this case we have 0.97 -551j as the inductor termination presented by the antenna. With ground losses normalized at 20 ohms and an inductor Q of

200, we have 20+2.76+.97 = 23.73 ohms of feedpoint resistance. Current is 2.05 amperes, and power radiated is 4.1 watts. Power lost is 95.9 watts.

Efficiency is 4.1%, a 6.3dB increase over a base-loaded triangular current system with the same lossy ground. This system is 8dB down from the same "top-loaded" distribution using a perfect ground.

When the system has significant fixed losses, increasing radiation resistance four times by top loading provides a similar dividend in system efficiency. At the same time a substantial increase in coil Q provides only minimal change in field strength.

Current Through Coil

Related pages:

Inductor spice model

There has been some speculation that current is high only in the first few turns of a loading inductor. Radiation comes solely from charge acceleration or current over spatial (in line) distance.

If any loading inductor shows substantial decrease in current over the length of the inductor, it is an absolute certainty that the inductor is poorly designed and that the system above the loading inductor is not contributing to system efficiency. The reason for this is very simple and straight forward. Any two-terminal component (even considering wire as a "component" applies) MUST have equal charges flowing into and out of each terminal. Voltages to other reference points can be different, but for every charge moving into one terminal a like number of charges MUST move out of the other terminal. Radiation, induction fields, and loss resistances have no influence on this rule.

In order to have any change in current, there must be an additional path or paths for charges. This path can be through leakage resistances, or through fictitious currents called displacement currents. Whatever the path, the total charge movements must be reconcilable. We simply can not have current "disappear".

The normal path upsetting "unbalancing" current into and out of each terminal in an inductor is provided by displacement currents through electric fields. As with any system, the amount of current flow is proportional to potential difference and impedance of the path. In order to shunt a substantial current out of an inductor, the potential difference between the ends of the path has to be high compared to the impedance of the path. The impedance of the stray path must also be reasonably low compared to the normal desired path.

Current diversion is problematic in very large inductors operated at (or very near) internal self resonance, when the self resonance is what we typically refer to as a "series-resonant" condition. This condition is common in plate chokes used in vacuum tube power amplifiers, where the system operates over many octaves of frequency range.

"Series resonances" inside components occur when distributed inductance forms a pair (or multiples of pairs) of "L" networks. The large series inductance from each end of a winding reacts with the small stray capacitance at the center, and forms a very high impedance transformation L network. The electrical potential at the center of the system becomes extremely high, and even the smallest amount of capacitance to surrounding objects will carry a substantial displacement current. The large displacement currents cause the terminal impedances to drop, and allow considerable current to concentrate in small areas of the component. At the same time, considerable voltage can be present. The normal result is arcing or destruction of the component, or failure of the system depending on the choke to operate.

Series resonance always occurs at a frequency higher than the self ***parallel*** resonant frequency of the component. A loading coil operating under such conditions would be required to have serious design errors to fall into this category, since the end termination capacitances should always be substantially higher than stray capacitance throughout the component. Failure to follow this rule would result in needless loss and reduced SWR bandwidth in an antenna.

The speculation or supposition that the first few turns of a loading coil carry most of the current is clearly untrue. In order to shunt current off, high series impedances would have to exist along with high stray shunting capacitance to areas removed from the radiator. Additionally, the remaining coil area connected to the top area of the antenna above the loading coil would have to present a high impedance to the area where current reduction occurs. This would never be the case, unless the top area of the antenna and loading coil are not resonant near the operating frequency.

A reasonable test for proper inductor and system design would be to remove the antenna above the loading coil, measuring system resonance. If resonance does not change substantially, the area above the coil is not correctly terminating the system. First-order self resonance of the inductor (parallel resonance), when removed from the system, should also be far above the operating frequency of the system. If self-resonance comes within three or four times the operating frequency range, the loading coil almost certainly will have needless performance shortfalls.

Conclusion

A normally functioning inductor has essentially equal currents throughout the inductor, loading coils are no exception. Any current difference requires a substantial current flow through undesired stray capacitances or leakage currents. Neither radiation or induction could change this, it is a basic rule of circuitry.

In a reasonably well-designed system, current into and out of the loading inductor should be substantially equal. Differences in current would indicate excessive and problematic undesired stray capacitance in the loading system design, or measurement errors.

Reduced sensitivity to coil Q is primarily a function of additional losses in the system, not reduction of current through the coil.

This page has had **visits since February 11, 2004.**

Independent Measurements

Tom,

Thanks very much for sending the copy of the posting. I'd appreciate it very much if you'd post the following on the same reflector for me:

It was brought to my attention that the following statement was recently posted on this reflector by Yuri:

"I will leave it here, as the rest of it. W7EL, author of EZNEC measured toroid coil and found that it HAS different current at its ends, roughly proportional to the part of antenna that it replaces."

This is not true. I assume the statement was made due to careless reading of the postings I made on rec.radio.amateur.antenna, rather than a deliberate distortion, so I'll very briefly describe the results again here. I made measurements of the current into and out of two different toroids in series with an antenna at its base. The first measurement was made using a 33 foot vertical with seven ground radials, whose feedpoint impedance measured 35 - j185 ohms at 3.8 MHz, and a toroidal inductor with reactance of 193 ohms and Q a bit over 300. I found that the difference in current between input and output of the inductor was 3.1% in magnitude and with no measurable phase shift, despite the short antenna. The 3.1% current difference between input and output can be explained simply by 6.8 pF of stray capacitance between the coil output and ground and/or current probe, or a bit greater value distributed over the coil. I repeated the measurement on the bench, with a series resistor and capacitor in place of the antenna, and measured 2.3% difference between input and output current. This is within measurement error of being the same as when connected to the antenna. (One would expect the stray C to be a bit different, also.) So in the one case I did the measurement with the inductor "replacing" a very significant

part of an antenna and in the other with the inductor replacing no antenna at all (or an entire antenna, if you choose), and got essentially the same result. How this can be interpreted as my having drawn the conclusion stated by Yuri is beyond me.

Unfortunately, the fact that the antenna is spaced only about 1/4" from a 4 foot mounting pipe (which altered its input impedance), in addition to the abbreviated ground system, left a great deal of wiggle room for people proposing alternate theories, so I did a second test with a more idealized antenna. For this test I constructed a vertical antenna that was 33 feet high, made of #16 insulated wire, and strung 23 radials out on the surface of the wet ground. The feedpoint impedance of this antenna at 3.8 MHz was measured as 15.8 - j437 ohms. This is close to the theoretical impedance, assuming about 8 ohms of ground loss resistance. The inductor had a reactance of 387 ohms. As described in my posting on rraa of November 11, the inductor "replaces" about 33 electrical degrees of the antenna.

The result from the second test was a current difference of 5.4%, again with no measurable phase shift. And again, this small difference can be explained by about the same amount of stray capacitance. It's nowhere near the 16+% that the "cosine rule" (that the output current equals the cosine of the "replaced" antenna) proposed by Yuri and others would predict. In no way did my measurement validate his theory -- 5.4% isn't even "roughly" more than 16%.

Frankly, I'm a bit embarrassed to have bothered to make these measurements at all -- it's a lot like making careful measurements to validate Ohm's law in order to refute someone's measurements that "prove" it wrong. As it is, it's turned out to be even worse than just a waste of time, since the results are now being distorted to support the very theory they clearly refute. I hope this will clarify just what measurements I made and what they showed.

For anyone who can stomach the waffling, backpedaling, and insults which characterized the original discussion, look up the thread "Re: Current in antenna loading coils controversy" and its variants in the google archives of rec.radio.amateur.antenna. Postings describing my measurements were made between November 8 and 11, 2003. If you read the thread, particularly the reaction to my posting of the measurements, you'll see why I've given up on that newsgroup as a forum for rational communication.

Many thanks to Tom, W8JI for bringing the misleading quote to my

attention and for posting this for me.

Roy Lewallen, W7EL

Roy's Measurements

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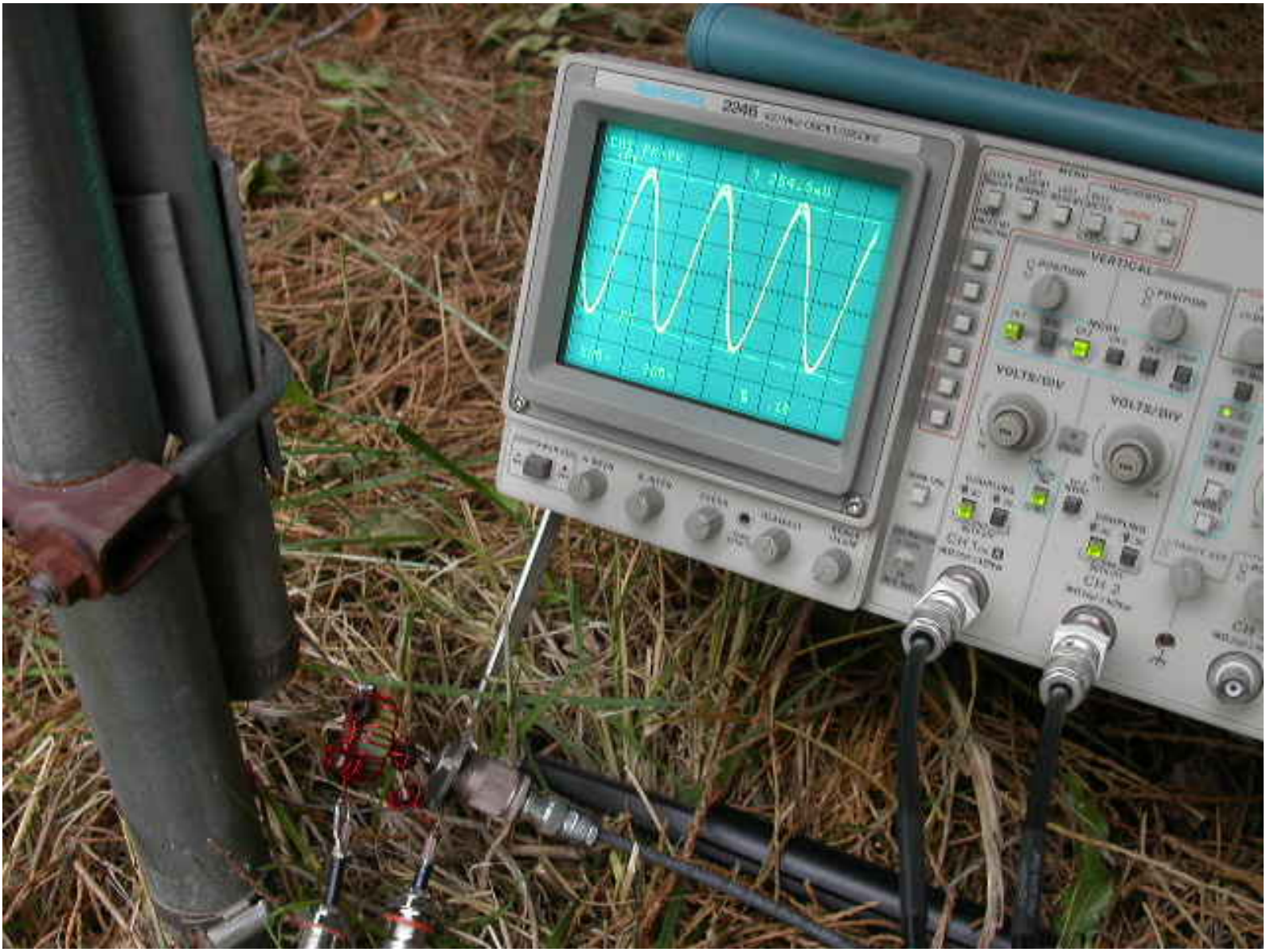
W7EL's Antenna current measurements:



This is the overall measurement setup described in the rec.radio.amateur.antenna newsgroup on and about November 9, 2003, showing both traces on the oscilloscope. Signal input is from an HP 8640B signal generator.

This is a closeup of the inductor and current probes. The upper pipe is the vertical antenna. The one to its left is the mounting, which extends about four feet above ground. The two are spaced 1/4" apart. The yellow toroid is the inductor being measured. The smaller black cores are the current probes. One core of each probe is a transformer, with the wire carrying the current to be measured passed through the core's hole. The second core is a common mode choke or "current balun", which serves two functions. One is to insure that no significant current flows on the outside of the coax cables connecting the probes to the oscilloscope. The second is to isolate the probes and reduce any possible capacitive coupling that might disturb the measurement. The probe outputs are terminated with 50 ohms at the scope.

Inductor operation

The following is a copy of RRAA postings by Roy Lewallen W7EL about inductors. I think the technical portion of these posts are useful for understanding current in loading coils, and how we need to cross check theories properly. I've included all of the text, even though much text is not technical. This really is educational, so read it if you have time!!!

I think the goal should be education, not conversation, so I enlarged important text size and changed color to blue in what I think are important areas.

Comments of mine are in RED.

W8JI Tom

The following are verbatim copies of postings made on rec.radio.amateur.antenna, in the thread "Re: Current in antenna loading coils controversy" (first posting), "Re: Current in antenna loading coils controversy - new measurement" (last posting), and "Re: Current in antenna loading coils controversy (long)" (remainder of postings). I and others made many additional postings on the topic, which should be available at http://www.groups.google.com.

Roy Lewallen

January 11, 2004
Posted on 11-4-03:
No, I will make one more comment. After a bit of reflection, I think this might be at the core of some people's problem in envisioning a lumped inductor.
When a current flows into an inductor, it doesn't go round and round and round the turns, taking its time to get to the other end. An inductor wound with 100 feet of wire behaves nothing like a 100 foot wire. Why? It's because when the current begins flowing, it creates a magnetic field. This field couples to, or links with, the other turns. The portion of the field from one turn that links with the others is the measurable quantity called the coefficient of coupling. For a good HF toroid, it's commonly 99% or better; solenoids are lower, and vary with aspect ratio. The field from the input turn creates a voltage all along the wire in the other turns which, in turn, produce an output current (presuming there's a load to sustain current flow). Consequently, the current at the input appears nearly instantaneously at the output. Those who are physics oriented can have lots of fun, I'm sure, debating just how long it takes. The field travels at near the speed of light, but the ability of the current to change rapidly is limited by other factors.

So please flush your minds of the image of current whirling around the coil, turn by turn, wending its way from one end to the other. It doesn't work at all like that. The coupling of fields from turn to turn or region to region is what brings about the property of inductance in the first place.

That was a great explanation of how an inductor works. I often learn new or better ways to view things when reading what Roy writes!

Radiation is another issue, and provides a path for current, via displacement current, to free space. (I can see it now in Weekly World News: WORLD FILLING WITH COULOMBS! DISASTER LOOMS!) For a component to fit the lumped element model, radiation has to be negligible. And, for the same reason, it can't be allowed to interact with external fields as a receiver, either.

This is very fundamental stuff. You can find a lot more about the topic in any elementary circuit analysis or physics text. If you don't believe what you read there, just killfile my postings -- you won't believe me, either, and reading what I post will be a waste of time for both of us.

Real inductors, of course, are neither zero length nor do they have a perfect coefficient of coupling. And they do radiate. The essence of engineering is to understand the principles well enough to

realize which imperfections are important enough to affect the outcome in a particular situation. We simplify the problem by putting aside the inconsequential effects, but don't oversimplify by ignoring factors that are important for the job at hand. Those who insist on using only the simplest model for all applications will often get invalid results. And those who use only the most complex model for all applications (as is often done in computer circuit modeling), often lose track of what's really going on -- they become good analysts but poor designers. I've seen people capable of only those approaches struggle, and fail, to become competent design engineers.

And with that, I'm outta here. Hope my postings have been helpful.

Posted 11-8-03:
Here are some preliminary details about the inductor current measurement I made.

My antenna isn't nearly as ideal as the one Yuri described. (But if my results are different from the ones reported at the web site Yuri referenced, I'll be eager to hear why.) It's about 33 feet high, and has only 7 buried radials. The feedpoint impedance indicates a loss of about 25 ohms at 7 MHz, so I'd expect it to be a bit more at 3.8. It's bolted to a galvanized fence line post which protrudes nearly four feet from the ground, with spacing between the antenna and the post of about 1/4". This mounting has only a minor effect on the feedpoint impedance at 7 MHz, which is the antenna's intended frequency of use. It's quite profound at 3.8 MHz, though. The expected 370 or so ohms of capacitive reactance is transformed to 185, while the feedpoint R is 35 ohms, not far from the expected value. So the overall feedpoint Z is 35 - j185 ohms at 3.8 MHz, measured with a GR 1606A impedance meter. (I found that my MFJ 269 was about right with the X, but measured R as zero -- apparently the combination of low frequency and large X is a problem for it in resolving the R.) So I built an inductor with measured impedance of 0.6 + j193 ohms. It's 26 turns on a T-106-6 toroid core. Q is a bit over 300. This was placed in series at the antenna feedpoint.

For current measurements, I made two identical current probes. Each one consists of 10 turns wound on an FT-37-73B ferrite core. The two leads from the winding are twisted and wound in bifilar fashion on another FT-37-73B core, 10 turns. This is then connected to an oscilloscope input via a two-foot (approx.) piece of RG-58. A 50 ohm termination is also at the scope input. This gives the probe a theoretical insertion impedance of 0.5 ohm. While making the measurements, I moved, grabbed, and re-oriented the coax cables, with no noticeable effect. This gave me confidence that the outsides of the coax weren't carrying any significant current.

One probe went to each channel of the scope. I left the two scope inputs in the cal position, put both probes on the wire at the input end of the inductor, and recorded the p-p values with the scope's digital measurement feature. Then I reversed the order of the probes and remeasured. I found a slight prejudice toward the probe closest to the source -- 1.2% in one ordering, and 2.1% in the other. Averaging the two channels, though, showed them to be the same within less than 1%. (Each probe was always connected to the same scope channel, so this is a test of the probe-scope channel combinations.)

Then I moved one probe to the output side of the inductor, and measured input and output current.

And I reversed the probe positions on inductor input and output. The ratio of output to input current in the two tests differed by only 1.4%.

When I encounter an astrologist, they invariably ask what "sign" I "am", then proceed to tell me how my personality meets their expectations. So what I do instead is to have them tell *me* what "sign" I "am" *first* -- which they should easily be able to do, based on my personality. Well, they don't find that to be fair, for some reason (although I certainly find it amusing). And so, I doubt if the following challenge will be regarded to be fair, for much the same reason. Those with alternative rules for solving circuit problems are challenged to predict what the ratio of output current to input current will be. I'm particularly targeting Cecil, and others who have bandied about a lot of pseudo-analysis about electrical length, reflections, and the like. And, Richard (Harrison), who said something like "an inductor without phase shift is like". . . I don't recall. . .hot dog without ketchup or something. Pull out your theories, and calculate it, like any competent engineer should be able to do. For cryin' out loud, it's a simple series circuit (except for Cecil, where it's some kind of distributed thing).

First post your answers, then I'll post the result of my measurements.
Roy Lewallen, W7EL

Posted on 11-11-03:
It sounds like the predictions are in. Among the several people who believe that the current out of a small inductor doesn't equal the current in, only Yuri was able to calculate a predicted value for the test, of 2.5 - 5% reduction in current at the output compared to the input, with a phase shift of about 18 degrees.

You see above Yuri "predicted" a current value based on a cosign formula that relates to antenna area. One way to test a theory is to change the conditions or parameters of the test, and see if the predictions apply in multiple cases or if it was just a "guess" or dumb-luck. This is ALWAYS a good idea. Watch what happened when Roy changed the antenna length and used the same formula!

What I measured was a 3.1% reduction in magnitude from input to output, with no discernible phase shift. The 3.1% is an average of two readings, with the input and output probes exchanged. The output was smaller than the input in both measurements, about 2% and 4%. So I believe there's a real difference between output and input current, although with the accuracy of my measurements, I only have reasonable confidence it's somewhere between 1 and 5%. I can resolve about 2-3 degrees of phase shift, though, and I couldn't discern any at all. (Yes, the scope trigger was from one channel, not

alternating.) So I have very high confidence that Yuri's prediction of 18 degrees is **incorrect.**

I don't subscribe to the notion that the current out of a very small inductor should be different than the current in due to some magical property it acquires when connected to an antenna. My working hypothesis is that the currrent difference I did see was due to stray capacitance, either from the probes or simply to the Earth and other objects. It would take an equivalent of 6.8 pF at the output of the coil (that is, between the coil output and the current probe) to get 3% reduction, and only about 1/3 that amount to see the minimum value of reduction of 1% I estimate was actually present.

Here is where Roy changed the test. Now the coil isn't in an antenna, and the "sine rule formula" should not apply. If the sine rule was correct, the current differential should have changed significantly.

I repeated the test on the bench, with a 36 ohm resistor in series with a 220 pF capacitor substituting for the antenna. The result was a 2.3% output:input reduction, again with no discernible phase shift. This is within the measurement error of being the same result. This is what should be expected -- except for unintentional coupling to the antenna's field, the inductor's environment is the same on the bench as at the antenna base, in these single frequency, steady state tests. (That also contradicts what some newsgroup participants have been claiming.)

So, although the small output:input current reduction was within Yuri's prediction, the phase shift certainly wasn't. If time permits, I'll make a more idealized antenna and repeat the measurements with a larger inductor at the base of a more reactive antenna. I'll predict in advance that if I double the amount of loading L, I'll approximately double the amount of current magnitude attenuation -- that is, to somewhere around 6%. That's what should be expected if the cause of the attenuation is stray C or a similar phenomenon.

I've added a picture to the http://eznec.com/rraa/Inductor_Current_Measurement.html page, showing the overall setup including the scope. It gives a little better perspective on the relative sizes of various objects.

Roy Lewallen, W7EL

Posted on 11-11-03:
Today's project was to construct and measure a more idealized antenna.

The antenna is 33 feet high, made of #16 insulated wire. I put out 23 radials on the surface of the

wet ground. Radials were of various lengths, most about 30 feet long. The feedpoint impedance of the antenna, measured with a GR bridge, was 15.8 - j437 ohms at 3.8 MHz. Allowing 3% lengthening effect for insulation, EZNEC says a lossless vertical of that height and diameter should have an input Z of 7.5 - j478. 8.3 ohms loss resistance is reasonable for that number of radials, and the somewhat lower than predicted reactance is likely due to the fact that the radial wires were grouped together as they came up a few inches to the antenna base, and not immediately coming in contact with the ground. That would add a bit of inductive reactance.

I wound an inductor on a T-106-6 core as before, but with more turns, for a measured Z of 1.3 + j387 ohms. After putting it in series with the antenna at the base, the base impedance measured 17.1 - j54 ohms. This is only 4 ohms from the expected reactance, and spot on the expected resistance, so measurements are consistent.

Analyzing verticals with EZNEC, made from #16 wire at 3.8 MHz, shows that:

-- An antenna 63.2' high is resonant.
-- An antenna 35.9' high has a feedpoint reactance of -j437 ohms.
-- An antenna 59.35' high has a feepoint reactance of -j54 ohms.

With a resonant height of 63.2', you could say that 63.2' is "90 electrical degrees" as far as the antenna is concerned. So you might say that my inductor has "replaced 33.4 electrical degrees" of the antenna.

Here again the test is modified to test Yuri's prediction.

Using Yuri's cosine rule, we should then expect the inductor output current to be cos(33.4 deg) times the input current, or 16.5% less. Also, we should expect to see those 33 degrees of "replaced antenna" as phase shift from the input to the output of the inductor. That is, the current change from the input to output of the inductor is the same as it would be for the portion of the antenna it "replaces". (I think Jim Kelley subscribes to this theory also, but I'm not sure.)

In contrast, conventional electrical circuit theory predicts no current difference between the input and output for a physically very small inductor with no radiation or stray coupling. I saw about 3% in the previous measurement, which I believe can be attributed to stray capacitance. So I predicted that we should see about twice that amount with the higher valued inductor used for this experiment (387 vs 192 ohms reactance). I didn't see

any measurable phase shift between input and output before, so I didn't expect to see it this time.
So for this test, there's quite a difference in predictions for output:input current --
**Yuri's method predicts a reduction of output current magnitude of 16.5% and a phase shift of 33 degrees.

**I predict around 6% magnitude reduction (due to stray C) and no measurable phase shift (less than 2 or 3 degrees).

I have very high confidence that my measurements are good enough to resolve the difference between these two possibilities.
Would anyone care to comment before I post the measurement results? And, Yuri, please correct me if I've misinterpreted your theory.
Roy Lewallen, W7EL

Here are the test results, still using the same formula. If the results of the Cos theory and formula are correct, the results would be 16.5% current reduction and 33 degrees phase shift. This shows why it is important to test new formulas in a variety of applications.

Posted on 11-11-03:
Ok,
For anyone who cares, the magnitude of the current out of the inductor in the later test measured 5.4% less than the current in. No phase shift was discernible. An analytical person could build on this information to investigate the properties of longer inductors placed elsewhere in the antenna.

Thank you for the comments, Cecil, Yuri, Richards, Art, and others. I've learned a good lesson from this -- that this isn't an appropriate forum or appropriate audience for the sort of quantitative analysis and reasoning I'm familiar and comfortable with. And that the considerable time and effort required to make careful measurements is really of very little benefit -- certainly not anywhere near enough to justify it.
With a great sigh of relief from everyone, I'm sure, I'll now turn this thread back over to Yuri, Cecil, et al.
My apologies to everyone for taking up so much bandwidth.
73,
Roy Lewallen, W7EL

So there we have it. The Cosine theory that states an inductor replaces part of

the antenna and has the same or similar phase shift and current taper as that part of the antenna proved false once the system was perturbed. It is quite typical for a formula that is incorrect to work at one point just by sheer luck.

When we have a theory, the test is to apply that theory to various systems and see if it works in all cases. If it does not work in all cases, the theory is incorrect.

While I agree it should not be necessary to prove conventional circuit theory, I think Roy's test (and my test) might shed some useful insight into how short and heavily loaded antennas actually work. A secondary benefit is we see how an incorrect idea or concept can actually work once in a while just by sheer luck, and why we need peer review.

Many people have the misplaced notion a wire in an inductor replaces a certain "electrical length" of the antenna. That isn't true. A loading inductor simply cancels reactance (corrects power factor). Current change across the inductor actually indicates a problem with the inductor, unless it is intentional.

Mobile Antenna Current Measurements at W8JI

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A sample of measurements above and below the loading coil with various antenna above the coil (current as percent of reference) follow:

	Toroid with hat	Small 2x2" coil with 24" hat up 24"	Long 12x3" coil with 24" hat up 24"	Long 12x3"coil with 6' whip
Current below	100%	100	100	100
Current above	100	94.4	73	76%

	Toroid with whip	long 12x1.5" coil with 6' whip	long 12x1.5" coil with 24" hat	Small 2x2" coil with 6' whip
Current below	100%	100	100	100
Current above coil	100%	79%	75%	96%
Current in whip 1ft above top of coil		73%		

The most revealing thing was how noticeably small changes in stray capacitance near the middle and top of the loading coil affect current distribution. It was quite evident hanging a large meter on each end of the coil would greatly perturb the system.

Mobile Antenna FS Comparisons

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The hat used in these tests was the DX Engineering Hot Rodz system.

Type	Height Base	Top Length	Hat dia	Field Strength	Frequency	Relative FS	
Reference				-18.5			best tested antenna
Large air	36"	84"	0	-18.50	7.2	0.00	3"dia #12
Large Air	36'	27"	47"	-18.70	7.2	-0.20	3" dia #12 47" hat
Small air	36"	27"	47"	-19.90	7.2	-1.40	1.5"dia #16 47" hat
TarHeel	42"	27"	47"	-21.00	7.2	-2.50	Tar with 47" dia
TarHeel	43"	84"	0	-21.10	7.2	-2.60	Tar with 7' whip
RM-20	36"	27"	52"	-21.20	7.2	-2.70	Hustler RM-20 on 40m
TarHeel	43"	27"	23"	-22.00	7.2	-3.50	Tar with 23" dia
TarHeel	61"	84"	0	-19.80	3.5		
TarHeel	59"	27"	47"	-19.70	3.5		47" DXE

The most striking things are:

- **The range between the best and worse system was only 3.5dB**
- **A large hat at the TOP of the antenna allows considerable shortening with no loss in efficiency**
- **For a given height antenna, efficiency is increased with a hat at the top**

Mobile and Loaded Antenna

Inductor Spice Model

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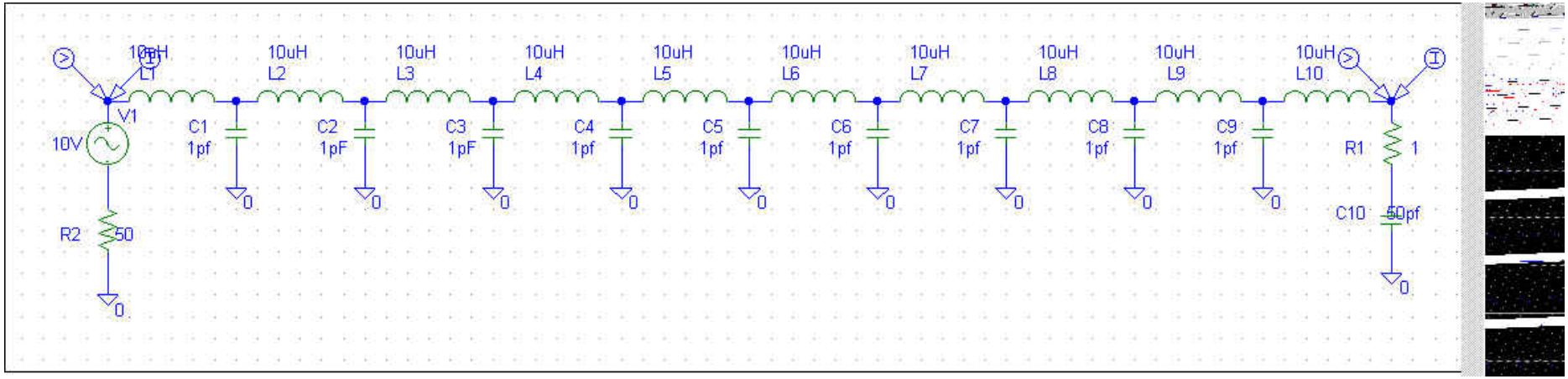
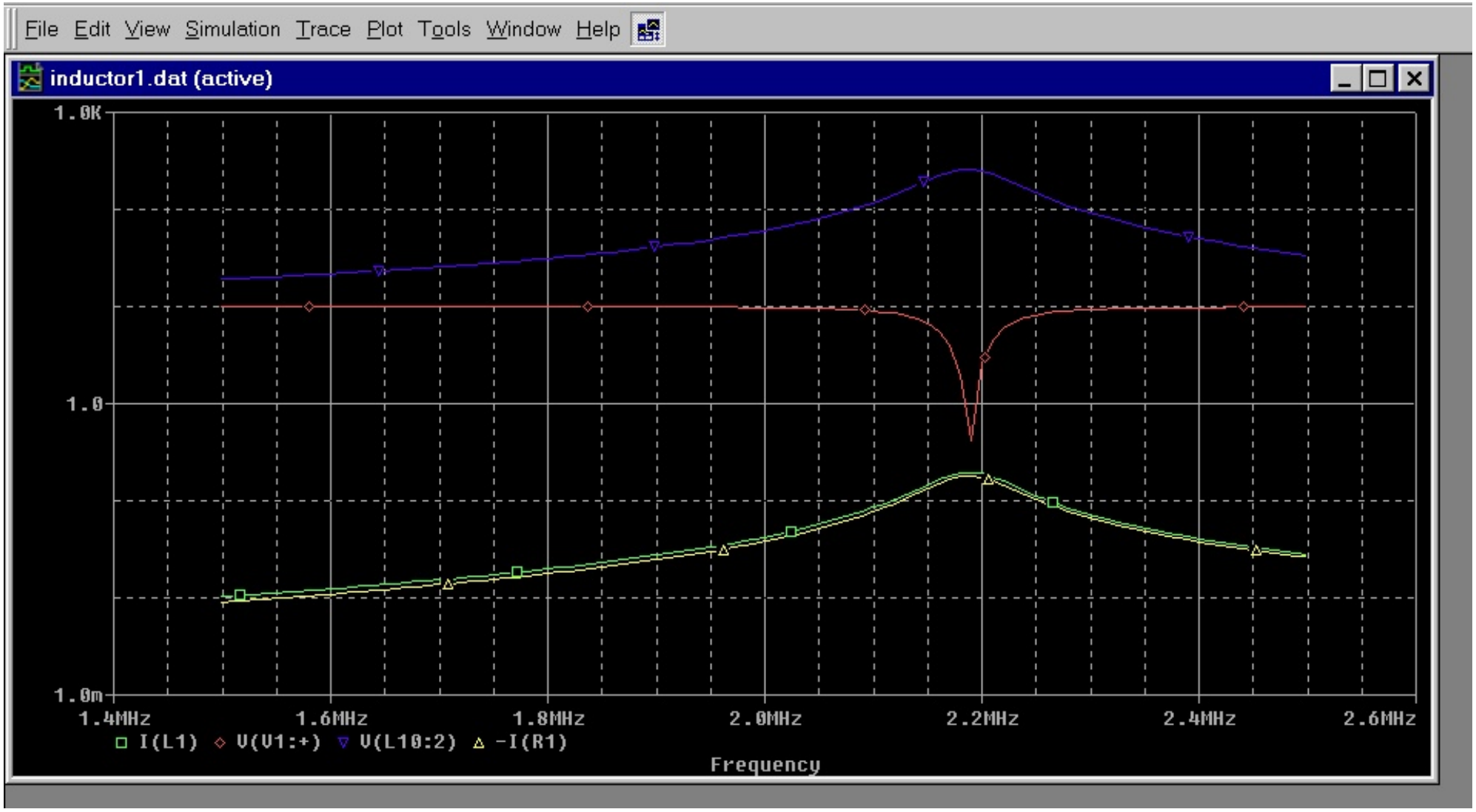
Spice model of inductor including stray capacitances:



Simulation.



Red=input voltage

Green=input current

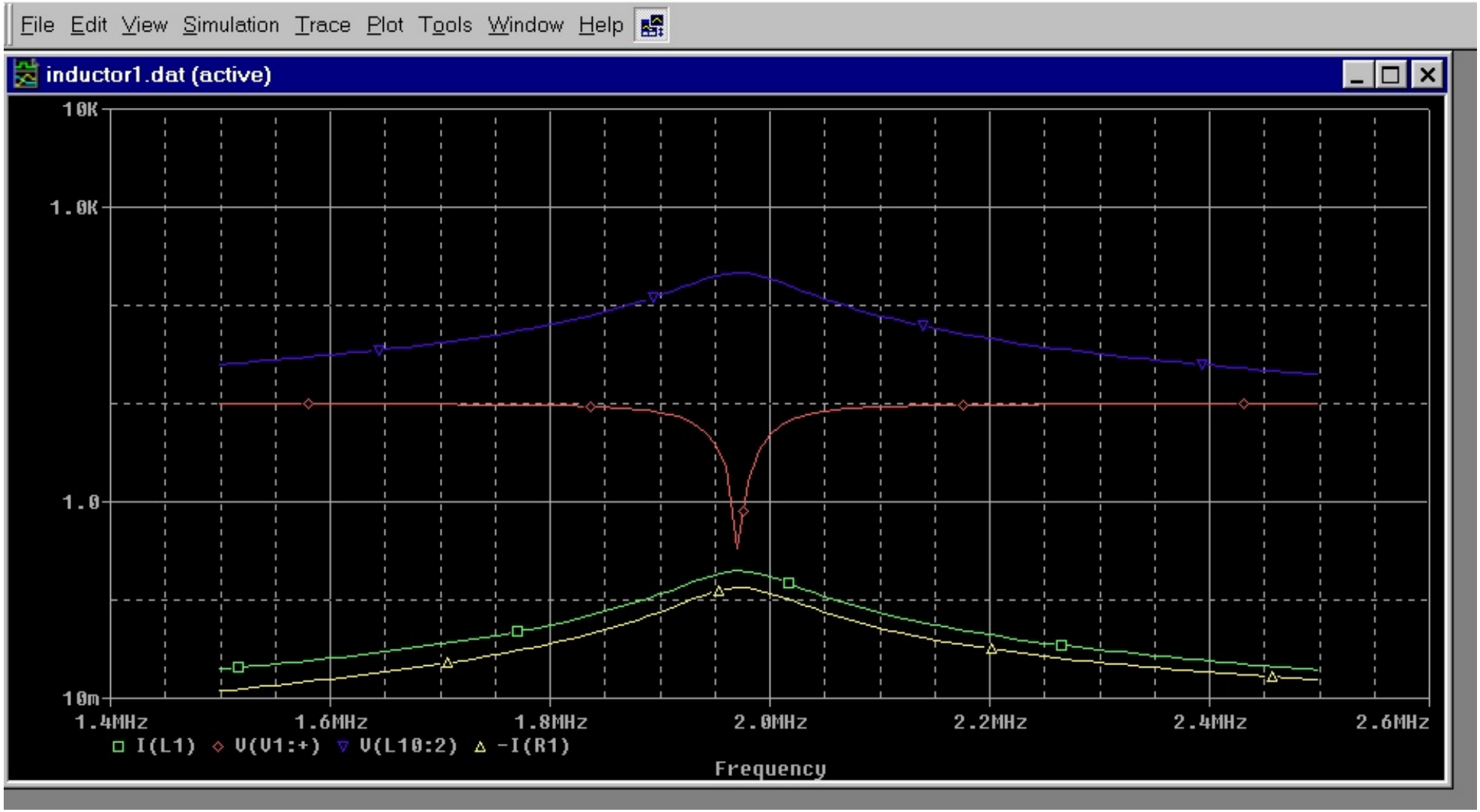
Blue=load voltage

Yellow=load current

This system (including reasonable leakage capacitance) representing a 50 ohm source driving a 100uH inductor with stray capacitance of 9pF and a 50pF load (small

whip), shows input and output currents are essentially equal, while output voltage greatly increases at resonance.

In the following SPICE plot, increasing stray capacitance shows expected departure from equal input and output currents. Distributed capacitance to ground is now at 45pF, nearly equal to the 50pF antenna capacitance:



Again despite stray capacitance nearly equaling load capacitance, current levels at input and output are reasonably similar:

Green=source current

Yellow=load current

Red=source voltage

Blue=load voltage

If we multiply output voltage times current, the apparent power is many dozens of times the input power. The reason is simple, phase relationship between voltage and current along the inductor changes even while current remains essentially uniform.

Although casual experimenters often assume a loading coil "replaces" the missing length in an antenna and has sinking voltage with rising current, this is not the case unless stray capacitance is extreme or the inductor is incorrectly and inefficiently operated at or near self-(series) resonance. The loading inductor really just corrects power factor, and bringing voltage and current into phase at resonance. It does replace missing length by simulating an antenna.

Linear loading is no different, it simply behaves like a very lossy low-Q coil and does not add radiation to the system. Unless of course we are talking about heat radiation.

Loading Inductors

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Optimizing Inductors

While intended mainly for antenna loading coils, the text below also applies to other resonant systems, such as amplifier tank circuits.

Related pages:

Inductor spice model

Mobile and loaded verticals

Optimum form factor of any inductor depends heavily on the actual end use of the component. Current through and voltage across an inductor as well as the required reactance influences both materials and form factor of an inductor. In high impedance systems, optimum form factor becomes longer with a smaller diameter. Even the best insulation materials will have a deleterious effect on component Q when impedances are high, and dielectric constant often becomes more important than dissipation factor in insulating materials. With low impedances, optimum form factors become more square and insulation has a much less noticeable effect.

Weight, size, and cost often require use of less-than-ideal materials and construction, but careful design generally results in a compromise that doesn't noticeably compromise system performance.

As a further complication, very simple systems might work quite differently than we intuitively believe. Many experimenters fail to consider what the inductor does in the system, and how the inductor behaves internally. For example many people assume, with a fixed amount of applied power, as voltage increases current must proportionally decrease. This isn't true in reactive systems, where voltage and current are not exactly in-phase or 180-degrees out-of-phase.

It is relatively common for programs and formulas to greatly overestimate maximum obtainable inductor Q. Confirming measurements are often flawed, either being made far from the operating frequency or with inadequate methods or equipment. This is especially true in high impedance high frequency systems.

Worse of all, antenna manufacturers and builders tend to seriously understate or estimate loss in other forms of loading, specifically linear-loading.

There are five common errors we should avoid:

- Building for excessive Q, when the reduced ESR will not noticeably improve system efficiency
- In antennas, considering inductor ESR directly as a portion of loss resistance at the point where radiation resistance is taken, rather than normalizing ESR to the point where radiation resistance is taken
- Believing programs or articles predicting Q's in the range of 1000 or more for inductors
- Thinking one optimum form factor (L to D ratio) always provides optimum performance
- Misapplying radiation or loss resistance formulas
- Believing claims that loading reactance obtained from stub or linear-loading provides lower loss than well-designed lumped loading

Before doing anything with information in this article or any other article related to loss and efficiency in antennas, please read the radiation resistance article on this site!

Range of Inductor Form Factor

There are two critical form factor dimensions, diameter and length. The ratio of diameter to length has two limits. The first limit occurs when the inductor occupies only one wire diameter as the length. The other limit occurs when the inductor is one wire diameter in diameter. The first condition would be met by a single layer pancake coil, the second by a linear conductor such as the inductance of a single wire transmission line. Optimum form factor occurs between these two extremes, and varies with the exact application.

Length-to-diameter ratio is important for two reasons:

- Shorter lengths and larger diameters increase capacitance across the inductor. Capacitance across any inductor carrying time-varying current increases circulating currents in the inductor, increasing loss while simultaneously reducing system bandwidth.
- Longer lengths and smaller coil diameters reduce mutual coupling between turns and increase leakage flux. This results in use of increased conductor length for a given inductance, increasing wire resistance.

These two situations are obviously in direct conflict, a balance must be achieved. Optimum balance between conflicting L/D effects listed above depend heavily on external circuit capacitance and operating frequency.

There is actually only one nearly constant parameter in design of high-Q RF solenoid inductors,

turn-to-turn spacing. Optimum turn-to-turn spacing occurs when the spacing or gap between turns is about the same diameter as the wire. If the turn-to-turn gap is filled (even partially) with insulation, optimum ***conductor spacing*** increases.

For the purposes of this article, the following terms are used:

- D=diameter
- d=turn diameter
- L=coil length

As a general rule, Q in a RF inductor peaks with a form factor (L/D) between 1 and 4.

The size and shape of the conductor used in the coil sets the optimum diameter, larger conductors require larger diameters.

Lower optimum L/D ratios (near unity) appear in systems where higher amounts of external capacitance load the system. Two examples would be amplifier tank circuits or large antennas with considerable loading capacitance beyond the coil. Another way to view this is by resonant frequencies. Form factor moves closer to 1:1 when an inductor is operated far below its natural self-resonant frequency.

Higher optimum L/D ratios (up to 4:1) occur when capacitance values external to the coil are reduced. Small mobile antennas without hats, especially top-loaded antennas, require longer form factors. Such systems operate the inductor closer to its self (parallel) resonant frequency.

The Reason Optimum Form Factors Vary

The underlying reason for change in optimum form factor with external circuit impedance rests almost entirely on inductor stray capacitance and mutual coupling between turns.

With high external capacitance, any reasonable amount of internal stray capacitance shunting the inductor causes a very small change in circulating current in the inductor. The external circuit, in effect, determines circulating currents. In this case, inductor Q is set mostly by flux leakage and conductor resistance in the inductor.

In such a system designers can place turns closer together, increasing mutual coupling or flux linkage from turn-to-turn. Since external capacitance causes most of the circulating currents, any increase in inductor distributed capacitance has little effect on the system. It becomes most important to reduce wire resistance by minimizing wire length. Dielectrics around the conductors have little effect on Q, because increases in capacitance caused by replacing air with a dielectric has little effect on the overall circulating currents.

As the system's external capacitance is reduced, circulating currents inside an inductor become increasingly influenced by stray capacitance. This includes capacitance within the inductor, as well as capacitance between the inductor and objects surrounding the inductor.

If inductor design or location is poor, and if system impedances are high, current can actually vary significantly along the length of an inductor. In properly designed systems, this will not occur.

When external capacitance is reduced, the coil ends must be increasingly separated from each other. The form factor chosen must reduce coil diameter while increasing length. In high impedance (reactance) systems, reducing capacitance improves component Q in spite of the resistance penalty resulting from increased conductor length required in long form factors.

We also must avoid using dielectrics near or in the inductor, especially any dielectric coating or between turns. Dielectrics other than air or vacuum, even low dissipation factor dielectrics, increase stray capacitance. Anything that increases capacitance will reduce component Q, and ALL dielectrics (other than air or vacuum) increase capacitance. The most noticeable effects in high reactance systems often come from dielectrics increasing capacitance, rather than actual dielectric losses!

The increase in loss can be directly proportional to the increase in capacitance, even when required turns are reduced. Low-loss Teflon or Polyethylene dielectrics can be nearly as detrimental as higher dissipation factor materials like fiberglass or Delrin.

Inductor Modeling Programs

Many inductor modeling programs fail to consider two important effects:

- They ignore capacitance across the inductor
- They ignore "current pushing" or bunching caused by strong magnetic fields

The first effect causes Q to peak well below the self-resonant frequency of the inductor. The second effect causes a decrease in Q as frequency is increased or as turns are brought closer together. The second effect occurs because current flows in a smaller and smaller cross section of conductor with increasing frequency.

If a model, prediction, or estimate does not show Q dropping drastically as first order (parallel) self-resonance is approached, the results almost certainly contain significant Q errors.

I've corresponded with some program writers who claim to have verifying measurements, and found their test equipment doesn't reliably operate (or operate at all) at the operating frequency of the inductors! Verifying inductor Q at frequencies far below the operating frequency in the model

does NOT provide any assurance the model or predictions are correct.

Optimum Q

There is a strong tendency to overkill the size of inductors, in an effort to reach unrealistic Q factors. Examples are commonly found in high-performance mobile antenna systems, where ground loss and other system resistances dominate the system. We often find high performance inductors with Q's in the several hundreds (at the upper practical limit of Q) and very low ESR's used in systems where overall loss resistances normalized to the feedpoint are very high.

Even though electrical problems are ***NOT*** created when using the highest possible Q, there is a point where the end improvement in signal level does not justify the physical size and cost to obtain "excessive" Q.

One example can be found in my Trap Measurements article, where differences in trap Rp (parallel resistance) when #10 AWG and copper tubing are compared, with unnoticeable changes in performance.

Another example appears in my mobile and loaded antennas article.

Inductor Placement in Antennas

The optimum location of an inductor varies with ground resistance and overall length of the antenna. Fortunately efficiency changes are smooth and gradual changes. Minor errors in placement generally do not result in noticeable efficiency changes.

Radiation Resistance and Mobile and Loaded Antennas articles on this site give some perspective of how load placement affects radiation resistance.

Q Ranges

The highest Q HF inductors I have measured, at least when operated away from self-resonance, are copper tubing coils and edge-wound inductors, such as those commonly used in high power tank coils. The highest Q I have measured in very large inductors of optimum form factor in the HF range has been near 1000.

Miniductor-type coils have a surprising amount of Q for the wire size, and maintain Q better as self-resonance is approached than larger coils.

These are the typical ranges of peak Q I have measured:

	HF Q Peak	Q at 80% of self-resonant freq
Copper Tubing Coils	600-1100	400-600
Edge-wound inductors	600-900	400-600
#8 miniductors	500-700	300-500
#12 miniductors	300-500	200-400
#16 miniductors	250-350	200-300
Large #2 mix iron on 1.8MHz	500-600	
enameled wire close-wound	200	100

As I measure inductors in the future, I'll include pictures and impedance data here.

Final Comments

We should keep the following in mind:

- Optimum form factor varies with application.
- Q peaks at some frequency significantly lower than self-resonant frequency, at self-resonance Q is zero (the coil appears as a pure resistance to any external circuit). Above that frequency inductor becomes the electrical equivalent of a low-Q capacitor.
- Linear Loading is really nothing other than a poor form-factor inductor. The radiation from the linear loading does NOT change the radiation resistance of the antenna except as the effective position of the load might change from the direction of fold. In all cases, a proper form-factor inductor would have less loss, and provide the same radiation resistance.
- Most inductor Q calculation programs overestimate Q.
- Any metal around an inductor decreases Q. Copper or steel, it often has nearly the same effect.
- Any dielectric (even low dissipation dielectrics) decreases Q because the dielectric increases shunt capacitance. This increases circulating currents. The effect is most pronounced as self-resonance is approached.

Radiation Resistance

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Related pages on Antennas , radiation and fields , mobile and short verticals

My 2004 Dayton Hamvention Power Point presentation on Small Verticals can be downloaded here... DAYTON 2004.

The main points when dealing with small antennas are:

There is no magic bullet or magic cure to make a small antenna act like a large one. It all comes down to current distribution over linear distance.

Small antennas require extraordinary care to obtain high efficiency.

How do we make a *small* antenna as efficient as possible?

- First, we make current as uniform as possible over the length of the antenna.
- Second, we use low loss loading such as optimum form (size, length, and diameter) loading coils.
- Third, we make the antenna as large and straight as possible in a line. We don't fold, bend, zigzag, or curve the antenna especially in the high current areas.
- We keep the high voltage points (the open ends) away form other things (like lossy earth), and the high current areas away from other large lossy conductors.
- Most important, we keep current as high as possible throughout the length of the antenna by using as much capacitance as possible at the antenna ends.

What this does is maximize radiation resistance (while at the same time minimizing loss). The text below explains how radiation resistance and loss interact.

Radiation Resistance

Radiation resistance is probably the most abused and misused term in antennas. The reason it is so often misused is the lack of clear definition. When a term has several nebulous meanings and uses, it is only natural that misuses will appear. The lack of a firm well-accepted definition allows meanings to slip from one application into another, where a totally erroneous conclusion can be drawn from what otherwise would be a good formula!

Common Uses

There are two commonly used "correct" meanings of radiation resistance and one totally incorrect use. The "correct" uses are:

- The resistive part of an antenna's feedpoint impedance that is caused by radiation from the antenna
- The total EM power radiated in all directions divided by the square of net current causing the radiation

Neither of the above definitions include loss resistances of any type! The moment loss resistance is included, we have the third commonly used (but totally useless) definition. This definition, which includes losses, must be considered incorrect because it is comprised of resistances that have nothing to do with radiation. The misused definition is:

- The real (or resistive) part of an antenna's feedpoint impedance including loss related resistances.

The correct name for the third "radiation resistance is actually the antenna ***feedpoint resistance***, not radiation resistance!

Of the above ***good*** definitions, the first definition is the most commonly abused through mistake. The second definition is an IRE definition (albeit a good one that never caught on). In every case I've seen, it is the second good definition that always provides the most direct and useful answer.

Examples of Misuse

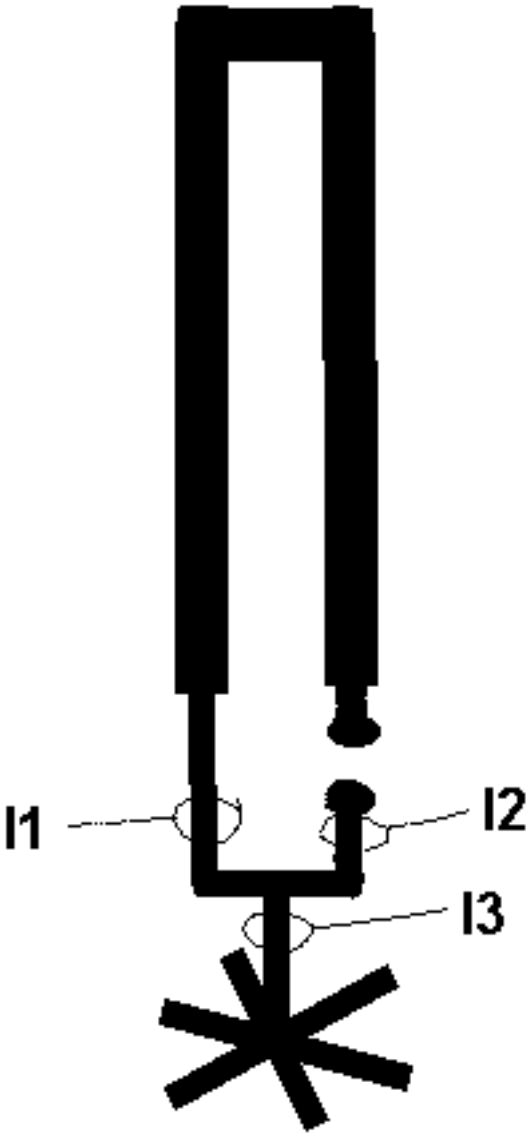
Folded Monopoles

Folded monopoles probably provide the best examples of misuse of the term radiation resistance. Quite often, in discussions of folded monopoles, claims are made that multiple drops increase the radiation resistance and lower losses. The justification for this incorrect claim is the folding raises

radiation resistance, and % eff = 100 * Rrad/(Rrad + Rloss) .

What folded monopole fanatics forget is that all losses must be normalized to the point where radiation resistance is taken, otherwise the efficiency formula won't mean a thing!

Let's look at what actually happens in a folded element, and use it to understand how the poor definition of radiation resistance causes the misunderstanding.



Consider the unipole above. Lets assume we short the open terminals, and feed it as a normal Marconi vertical with a feedpoint at the point where we measure I3. I3 is ALWAYS the vector sum or in-phase combination of currents I1 and I2. With continuity through each leg, I1 and I2 share all of the ground current. This happens regardless of where the feedpoint is located in the lower portions of the antenna.

With 1/4wl height and a reasonable element diameter, the radiation resistance (fed as a traditional monopole) would be about 36-ohms.

Assume ground loss, normalized to the point where we measure I3, could be represented by 14 ohms. Applying 500 watts would make current I3 equal 3.16 amperes. **Power loss in ground resistance would be I^2R, or 3.16^2 times 14, about 140 watts.** Feedpoint resistance would be 50 ohms. Feedpoint power, as a check, would be 3.16^2 times 50 ohms or 500 watts. With equal diameter legs, that current would divide and 1.58 amperes would flow in each upper leg.

Let's use the formula %eff = 100*Rrad/(Rrad+Rloss). We have 36/36+14 = .72 so the result is 72% efficiency, 28% loss. **28% loss times 500 watts is 140 watts in ground losses.** This matches the other method just above.

Opening the terminals and feeding as a folded unipole, half of the radiator current is in I1 while the other half is in I2. Current is halved to 1.58 amperes at the feedpoint and power remains the same. The feedpoint resistance now becomes 200 ohms. We can confirm this with I squared R, or 1.58^2 *200=500 watts. It all works out great so far!

Now let's MIS-use the same efficiency formula, like Orr does in his Handbook and others do in various places. We would have 200/200+14 = .9346 or 93.46 % efficiency.

We know we still have 3.16 amperes flowing as I3, and we know ground resistance is still 14 ohms (normalized to the point where I3 is measured). I-squared-R losses are 3.16^2 * 14 = 140 watts! We have exactly the same power loss.

Let's transform the ground loss value that was normalized at 14-ohms where I3 is measured to the feedpoint by the same impedance multiplication as the feed resistance, or 1:4. We'd now think ground resistance would be 4*14 = 56 ohms. 56 ohms of the feedpoint resistance is loss. Trying that same efficiency formula, we get:

144/144+56 = .72, or 72% efficiency!!! Now everything checks out fine.

The Common Mistake

Orr and others have used the first definition of radiation resistance, the portion of the terminal

resistance of the feedpoint responsible for radiation. Unfortunately ***they failed to normalize ground losses to the same point where the radiation resistance was taken!***

We can not use a formula that is based on everything being normalized to one point, unless we actually do that for every term in the formula!

There is no change in efficiency when the NET radiator current remains the same, and when ground current remains the same. It is pathological engineering to think otherwise.

Using The Second Definition

If we use the second IRE definition of radiation resistance, where the effective current causing radiation is compared to power radiated, we find nothing changes. A folded dipole or monopole has the same radiation resistance as a regular dipole or monopole the same size, and a small loop has the same radiation resistance regardless of turns.

The magic vanishes along with the incorrect definitions and perceptions.

You can read about this in textbooks. The "Antenna Engineering Handbook" by Jasik in 3-13, 19-3, and in other sections uses correct definitions and descriptions.

Quad's and other Loops

We find the same efficiency misconceptions in articles about small loops and large quads. Authors sometimes assume, incorrectly, radiation resistance changes in a favorable proportion to loss resistance as we make the feed impedance increase. What we really are doing is placing the feedpoint in series with a smaller portion of NET current causing radiation.

With a large full-size quad element the pattern under some conditions will slightly change, but efficiency remains basically the same. With a small loop antenna, losses can actually increase with more turns!

Terminated Folded Elements

Another abuse of radiation resistance is found in terminated antennas. Some manufacturers and authors claim a resistance can be inserted in series with one leg of a folded monopole or folded dipole, and the other leg fed. The usual arm-waving claim is the antenna isn't really resistor loaded, and that efficiency is very good because radiation resistance is high.

That claim is absolute nonsense!

A large terminated Rhombic is well-known to have poor efficiency. Rhombic gain is actually low compared to other antennas having the same sin/sin x antenna pattern, because Rhombic efficiency is generally less than 50%. At least half of the power is consumed in termination and ground losses below the antenna. The actual gain may be reasonably high compared to a dipole, but not to other efficient antennas with the same half-power beam width.

The typical manufacturing buzz-word is that terminated monopole and dipole antennas are "traveling wave antennas" and by some magic (that even large terminated Rhombic antennas can't achieve) have broad bandwidth and high efficiency.

A Rhombic focuses energy (that is not transformed into heat) into a narrow beam that has considerable gain, but if it sprayed the radiation around in a non-focused pattern, a regular dipole would win hands down. Throw a resistor on that dipole to smooth SWR variations, or on a vertical, and efficiency suffers.

I listened to a station on 75 meters 600 miles away testing a Sommer T-25 vertical. He was 30 over nine using a dipole, and dropped to S6 with the vertical. The European he was working reported a similar change. By removing the termination resistor and base-loading the same vertical, a local Ham gained almost 25dB on 80 meters!

When we abuse or misuse radiation resistance, we can invent all kinds of magical antennas. We can have CFA's, E-H antennas, terminated dipoles, small magnetic loops, and verticals with all sorts of magical claims. Few, if any, of the claims are ever correct. Any time we see a claim that efficiency changes a ***large amount*** because of feed method change, it should be a red flag.

Increasing Radiation Resistance

Radiation resistance, at least under the useful IRE definition, can be defined by the following formula:

$$Rrad := 160 \cdot \pi^2 \cdot \left(\frac{He}{\lambda}\right)^2$$

which would translate to:

$$Rrad := 1580 \cdot \left(\frac{He}{\lambda}\right)^2$$

Where *He* is the effective height center of accelerating charges that cause radiation. In other words,

He is the effective height, expressed in fractions of a wavelength, of the distributed common-mode current in the structure.

(Common-mode current is the vector sum of all currents, or the effective in-phase current at any point, or the current we would measure if we placed a giant clamp-on current probe around ALL of the conductors at that any given height.)

He and λ must both be in the same units, either given as degrees or decimal fractions of a wavelength.

As an example, a uniform current single conductor antenna has an actual physical height of 15.19 feet on 1.8 MHz, where one wavelength is 546.67 feet, the effective height is:

15.19/546.67= 0.0278 wl

Since charges are distributed evenly throughout the structure the full height is used. The effective height is .027 wl, the same as the physical height. The height in electrical degrees is .0278 * 360 = 10 degrees

We have a radiation resistance of:

$He := .0278 \qquad \lambda := 1$

$$Rrad := 160 \cdot \pi^2 \cdot \left(\frac{He}{\lambda}\right)^2$$

$$Rrad = 1.22$$

We can express this graphically in a chart, such as one found in the Antenna Engineering Handbook by Jasik:

Finding 10-degree height on the graph above, and following that line until we reach the crossing for

unity current ratio, we see the ~1.27 ohm radiation resistance is in agreement.

Notice that the number of vertical conductors does NOT enter into the equation! This is the absolute maximum possible radiation resistance we can obtain for a given radiator height.

Non-uniform Current

Radiation resistance is purely a function of the effective current distribution and height of the radiator, and is limited by height (spatial length)! Current throughout the antenna will not remain uniform if we reduce the size of the flat-top or hat.

Current will become zero at the very top with no hat, and 100% base loading. In this case, with no change in height, radiation resistance will be approximately 1/4th the value of the uniform current example. The result is exactly like a 50% reduction in effective element height.

If we follow the 10-degree line to the intersection point with 0 top current, we find radiation resistance to be around .32 ohms. 1.27 ohms, the radiation resistance for uniform current, becomes 1.27/4 or .3175 ohms.

If we stay on the uniform current line, we find that .3175 ohms would be the radiation resistance of a 5-degree monopole with uniform current.

Efficiency

It often helps to look at the extremes, so we can get a feel for the effect of changes.

Let's look at the poor ground extreme and assume we have system losses, normalized to the current maximum, that are many times the radiation resistance. This would be the case for a short 160 or 80 meter mobile antenna.

In such a system radiation resistance would dominate any change that would affect efficiency. Current distribution would mean everything to efficiency.

Assume we have a base loading coil, either good or poor, and a thin mobile whip above the coil. Efficiency would increase by a factor of approximately four times by installing a capacitance hat with several times the distributed capacitance of antenna conductors below the hat.

Moving the coil would have little or no effect on efficiency.

A six-foot antenna with a large hat would be electrically equal to a 12-foot antenna without a hat.

This is why very poor inductors used on antennas in mobile shootouts, with large hats, equal or beat very large high-Q coils in similar height antennas that do not have large hats. One case in mind was a Hamstick lash-up in a mobile. The Hamstick, a notoriously poor efficiency antenna for 75-meters, soundly trounced Bugcatcher antennas when a large hat was added to the Hamstick.

Moving the coil up on the antenna has the effect of making current below the coil uniform, but without a hat current above the coil is a triangular taper that reaches zero at the element tip. The effective height of the area above the coil is 50% of actual height.

If we add a large hat at the bottom of the whip, current in the whip is actually reduced! At the same time, we change nothing below the coil. The effect of adding a large hat below the whip is to reduce the effective height of the antenna, when considered as a percentage of physical height. Radiation resistance and efficiency is generally reduced by adding a hat just above a coil, even if the hat allows us to use a smaller coil!

Adding a large hat at either end of a coil also reduces coil Q, since a large portion of the hat capacitance directly shunts the inductor.

Conclusion

We can reach the following conclusions:

- Radiation resistance, or at least the useful definition of radiation resistance, is limited by spatial area (or height in the case of a vertical) any antenna occupies.
- Radiation resistance is maximized by making current as large as possible over the entire spatial area of the antenna.
- Surrounding objects generally reduce radiation resistance and efficiency, even when they are NOT resonant, because they reduce effective height! This includes dielectrics that increase capacitance of the antenna to ground, since any increase in capacitance appearing well below the top of an antenna reduces effective height.
- Radiation comes from charge acceleration, nothing else. The longer the linear spatial distance we move charges in, the fewer charges we need to move at any point for the same amount of EM radiation. This is just another way of saying radiation resistance is higher in physically longer structures, especially when they carry uniform current.
- Any antenna (including Linear Loading, Helical Loading, Folded Monopoles, Fractal shapes, CFA, E-H, and so on) claiming to increase radiation resistance beyond the limits outline above is based on misunderstandings or untrue distortions of basic antenna principles.

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Radiation and Fields

[Home]

Related pages Radiation Resistance

We tend to think of electric, magnetic, and radiation fields as physical "things"...much like an element would be in chemistry. This can cause problems!

One popular (but incorrect) assumption is we can combine, mix or blend fields into another field. After all.... we have electric fields...we have magnetic fields...surely we can mix them and create an electromagnetic field. This would have a distinct advantage, because we all know nearfield losses are higher than farfield losses. We know the farfield field ratio (or field impedance) is set by the media the wave propagates through. With reasonably dry air or vacuum, at normal radio frequencies, we know the impedance is 377 ohms. Why not just mix the two fields and create the third field without all that needless nearfield loss?

If, in our mind's eyes, we can mix fields the next logical step would be to conclude we can separate, filter, or sort them out whenever handy. If we have a problem with electrical interference, we can just eliminate the electric field. If we have a problem with field intensity (defined as volts-per-meter) causing problems with the neighbor's VCR, we can reduce the electric field and cure the problem. After all, it is electrical interference, isn't it?

We logically might assume, just by name, we can mix the various fields and make a new field. We logically might assume we can split the combined fields back apart if we don't like what is happening to one of them.

Unfortunately, none of the above is true! The above conceptual problems started at the very beginning, because we assumed fields were physical things like building blocks. They aren't.

Fields are really just mathematical descriptions of forces between charges. We have three simple conditions that create physical forces between charges. They are electric, magnetic, and electromagnetic (radiation) forces. They are all created by distinctly different physical actions in a system. We can't mix the various names for effects resulting from a certain physical condition and create a new cause! Fields describe effects, they don't create the causes of those effects!!!

Electric Field

Electric field describes a force created by uneven charge distribution. Nature wants charges to be evenly distributed, she can only take so much piling up of charges in one spot! The force between charges, caused by nature trying to balance or even the distribution of charges, is called an ***electric field***.

Uneven charge distribution goes hand-in-hand with a voltage difference between two physical points. We can obviously can have a difference in charge distribution in insulators as well as conductors. A comb, "charged" by running through our hair (if we have any left), can have an electric field. The force of this field can pick up tiny bits of

paper, as nature tries to equalize the charge distribution. The terminals of a battery have an electric field between them, and when we place a conductor in that field the charges try to equalize. Another example would be an antenna, where a voltage difference (uneven charge distribution) between two points creates an electric field.

Any difference in ***charge distribution***, whether the charges are "moving" or standing, causes a physical force. We name this effect an **electric field,** and describe it mathematically in volts over a given distance. This field (or force) decreases rapidly with distance.

Magnetic Field

Magnetic field describes a force created by moving charges. When charges are moving, they exert a force on all other charges around them. We call this effect the **magnetic field.**

One example of a magnetic field is a conductor carrying current. Perhaps it is a wire connected between two terminals of a battery. Another example would be a RF-current carrying conductor in an antenna.

Any movement of charges causes a magnetic field, and somewhere rooted in the creation of that magnetic field is an uneven distribution of charges causing an electric field! Once again, this field (or force) decreases rapidly with distance.

Electromagnetic Field

An **electromagnetic field** is created whenever charges are accelerated. Acceleration occurs whenever a charge changes direction or velocity. When a charge accelerates, all the other charges in the universe feel a force trying to make them move. The only thing that stops that force from going on forever is if another charge (or combination of charges) accelerate to create an equal but opposing force. We call the velocity at which this force ripples through the universe the *speed of light*.

One example of electromagnetic fields is in an AC current carrying conductor, like a power line. The time-varying voltage causes charges to move back and forth, and the change in velocity and direction causes the effect called electromagnetic radiation.

It is plain to see why our antennas have all three fields, and why we can communicate so well over large distances with low power. While the electric and magnetic induction fields drop off rapidly, the (initially) much weaker electromagnetic radiation field goes on until something cancels it. It is the radiation field that allows us to communicate, not the electric or magnetic induction fields!

Near the Antenna

Near any antenna fields are a complex mixture or " soup" of various effects from charges. Charges are almost always moving in multiple directions and over various distances, when viewed from distances close to the antenna. It isn't always easy to understand what actually is happening very near antennas (where the area of the antenna is large compared to the distance from which we observe the effects of charges). Near the antenna, pattern and field impedance is generally nothing like we might intuitively imagine!

It is the response in this area, generally within $\lambda/10$ distance from the antenna, that small "magnetic loop" and

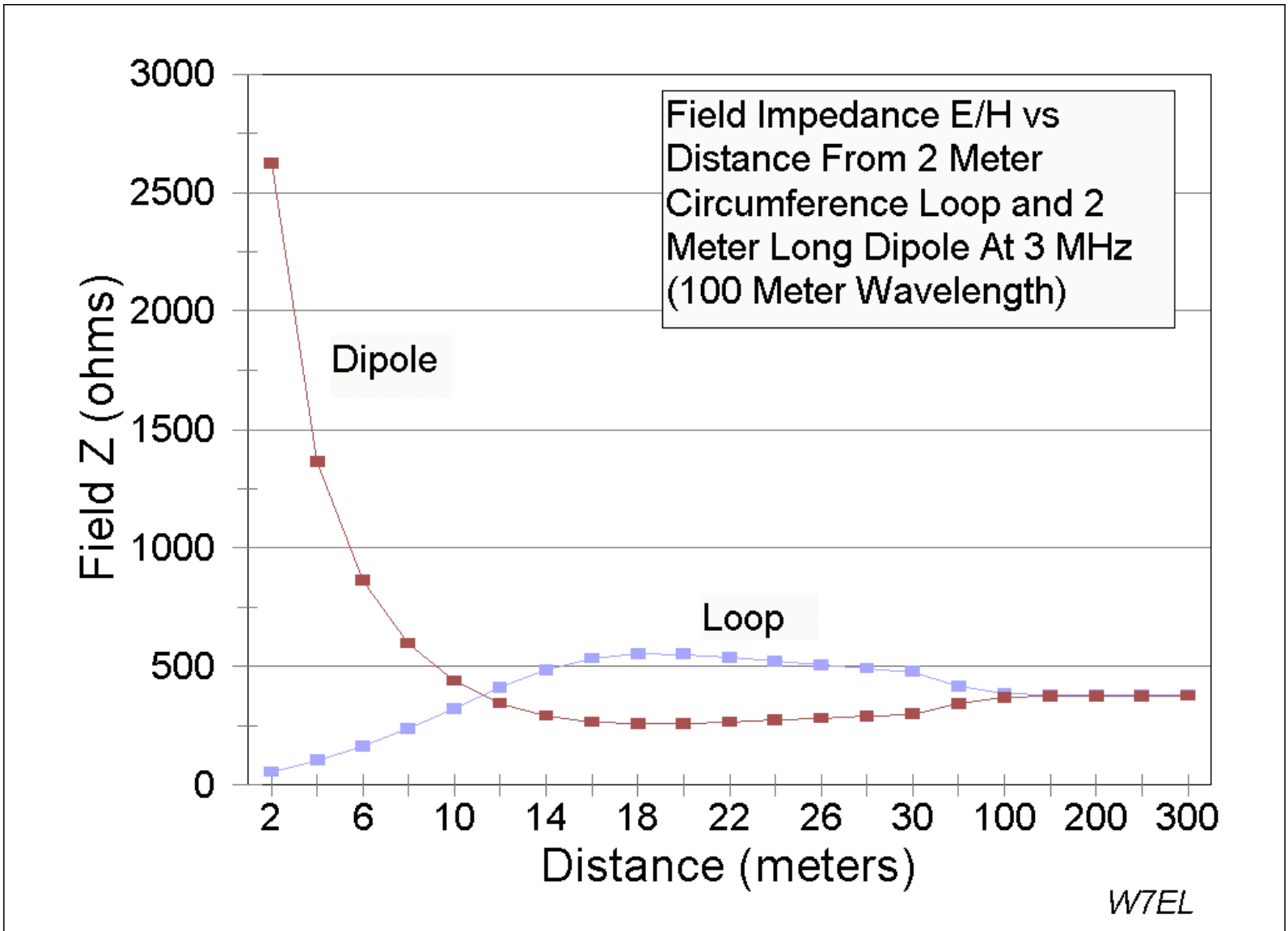
"electric dipole" antennas get their names.

Very close to the high-current area of a small loop antenna (but not near the open ends, because that is where the electric field dominates), the magnetic field dominates. Magnetic fields are mathematical descriptions of forces derived from moving charges, or current flow. This effect, when large compared to the electric field, is sometimes described by saying the "***field impedance***" is "***low***".

Conversely, near a small dipole or monopole with high voltage and little current, the electric field dominates. The largest force is from the very high open-end voltages, and very uneven charge distribution. We might say such an antenna has a " ***high field impedance***" in the area where the electric field dominates any forces cause by moving charges.

In all of these cases, if the antenna is electrically small, the dominant fields apply only within approximately λ/10 distance from the antenna!

As we move out further the weaker radiation field, because it is attenuated less with distance, starts to have a noticeable contribution to the charge forces. Because the phase of the fields (fields are a way of describing effects) is different at the antenna, the sum of the effects is different with distance. At some distance the low field impedance of a small loop becomes high, and the high field impedance of a small dipole becomes low!



W7EL

Since the distance of a wavelength in the above graph (thanks W7EL) is 100 meters, we can also considered the bottom scale as a percentage of a wavelength. We can see at about 11 percent of a wavelength (which would be about 50 feet on 160 meters), there is no difference in field impedance between a small loop and a small dipole. At distances beyond 50 on 160 meters, the loop actually has a ***higher*** field impedance than a dipole.

Losses

None of this has anything to do with losses being high or low in small antennas. Losses are directly related to the field density, and when we are close to any antenna the fields are very intense. Losses are not a field ratio problem, they are a field intensity related. In very small antennas, virtually ALL of the losses are related to reactance canceling and resistive losses in the antenna and any lossy media around the antenna!

We also must be mindful of the painful truth that we can not take either field to zero, or all radiation stops. By definition, radiation is an electromagnetic wave. We can't modify the field impedance of an antenna without changing the voltage and current distribution of the antenna.

Nearfield

The nearfield area is an area where the ultimate pattern is not fully formed, and where induction fields (from charge distribution and charge movement) have a noticeable effect.

It is possible, with large arrays of small elements, to be out of the induction field region but still find the area called the "nearfield" area or zone. Let's consider individual groups of elements as "cells", and the array a combination of small directional cells occupying a very large physical area.

Each cell has formed a radiation field. Depending on the size and type of radiator in each cell, the induction fields that charge distribution plays a role in may be attenuated so much as to be negligible, yet the radiation pattern of the entire array may not be totally formed.

This is the case with my phased Beverages and phased verticals, where the individual antennas making up the array are so distant that the effects of charge distribution (electric induction field, sometimes called the electrostatic field) or steady movement (considered at one infinitely brief instant of time, or magnetic induction field) have no effect. For example, at about 1 wavelength distance the electric and magnetic induction fields are negligible from either my circle of eight verticals or 780-foot Beverages, yet the pattern of the overall array established by the phasing of multiple cells is not fully formed. The pattern would only be fully formed several wavelengths from each array, where the distance between cells or elements is a small fraction of the viewing distance.

The total pattern of two 780-foot long Beverages spaced 350 feet apart is not fully formed even at distances of several thousand feet, yet nearfield induction effects are totally gone at much shorter distances. The field impedance is established, yet the antenna pattern is not.

The nearfield generally refers to or includes the area where "static" or induction fields still have a noticeable influence.

Fresnel Zone

The Fresnel (fre-nel, no "S" sound) zone is the area where pattern is still being formed. It may or may not include induction field areas.

Physically large arrays almost always have a physically large Fresnel zone. Even simple omni-verticals have a Fresnel zone extending out a few wavelengths. The field impedance may or may not have already been established in the Fresnel zone.

You may have heard about Fresnel zones during discussions of vertical antenna loss at low wave angles, or Fresnel lenses for lighthouses or other beacon lights.

Farfield

The farfield is the area where any changes in distance result in no noticeable change in pattern or field impedance. Losses are lower in the farfield area because field density is lower, not because we call it farfield.

Summary

There really isn't a clear distance where certain effects just abruptly stop. Transitions are smooth and gradual, because transitions are the result of distance....nothing else. These regions of effects are all distance related, and caused by the three distinct effects of charges moving, charge distribution, and charge acceleration.

Even if we could somehow make the field-impedance a certain value or electric and magnetic effects a ratio equal to farfield ratio, it would not eliminate or in the slightest amount reduce nearfield losses!

Losses are only related to the density of fields in any given media or environment, and nearfield or Fresnel zone losses decrease with distance because of the lower field intensity in any given lossy area, or wider cross-sectional area of lossy media carrying the same energy.

Larger antennas, in general, have larger boundary areas for electric and magnetic fields. With wider boundary areas, the system has less-concentrated fields and charge effects. Very small antennas of any type obviously have very concentrated fields, and the high field density or concentration of current or voltage in conductors are at the root of increased loss. The concentration of energy in a small area is the real reason losses are generally much higher in smaller antennas, and this has little or nothing to do with field impedance!

We can not mix fields and create a different field and make a small antenna become an "artificial large antenna", nor can we "filter" fields and remove the electric field to reduce electric noise.

This page has been visited

times since June 12, 2004

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Traps

[Home] [Up]

This article is from a talk I gave about antenna traps. It contains measurements of traps, performance of trap antennas using models of traps, and ideas on how to make trap antennas more efficient.

Try taking this Trap-Q test! Be honest.

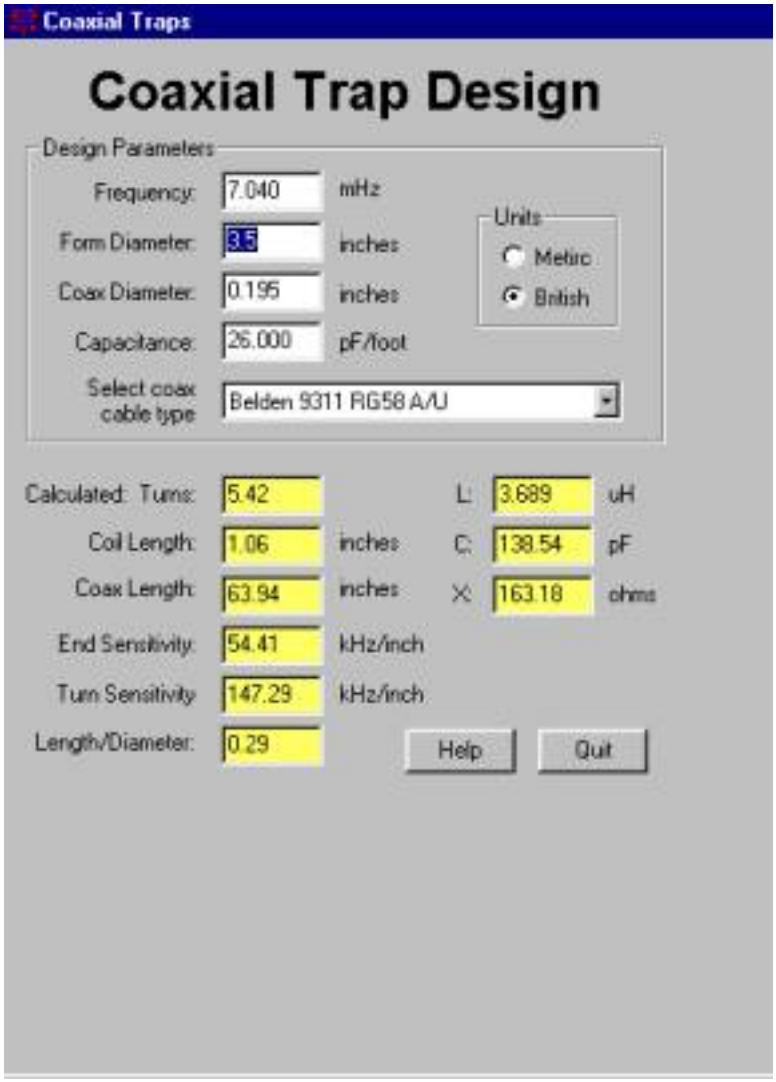
1.) Is it best to make the trap resonant close to the desired operating frequency?

2.) Does bandwidth decrease with increasing trap Q?

3.) Do traps create noticeable loss, perhaps one dB per trap typically?

4.) Does higher trap operating Q always mean lower loss?

Coaxial Trap Designer by VE6YP (Tony Fields)



This is a good program to get you in the ballpark with a trap design. It was available as freeware. (Unfortunately coaxial traps are relatively lossy on the trapped frequency compared to other types.)

The software is available at **http://members.shaw.ca./ve6yp/index.html**

7.04 MHz 3.5 inch diameter form RG58/U into the VE6YP program yields calculated values of:

Calculated	Actual Measurement
L= 3.689µH	3.116µH
C= 138.5 pF	164 pF
64 inches	59 inches

Using the program TLA by N6BV (from ARRL), we would estimate capacitance of a 59" RG-58/U cable as:

R .22 –j143.61 or about 157 pF (Q=650)

Measuring a real-world stub, capacitance was 164pF (Q=590).

While that Q seems high, remember a typical transmitting-type air-variable capacitor has a Q of several thousand!

Coaxial Trap Articles and Programs use capacitance/ft multiplied times length….

26 pF * 4.917 feet = 127.84 pF in trap program

C164 pF measured. This error, 36pF low from 164pF, occurs because the transmission line making up the "coaxial capacitor" is not actually treated as a transmission line in the modeling program.

Fortunately the error is in a useful direction, because we can shorten the cable! Coaxial capacitors are really open stubs, and should be treated that way once they are more than a few degrees long.

CONCLUSION: The difference between TLA and an actual measurement was around 4%. This is very close, but the result has significant difference from the coaxial trap program since it only considers pF per foot as the capacitance. A longer cable (in fraction of a wavelength) results in greater error by using pF per foot. The error comes because a coaxial cable capacitor is really a stub, NOT a pure capacitor!!

Trap Measurements (at resonance)

Type	F MHz	R parallel	X
Coax RG-58	7.034	17,800	0
UT-141-75 semi-rigid	7.045	45,330	0
100pF 7.5kV & #12 wire	7.040	99,850	0
60pF 15 kV & #10 wire	7.040	250,000	0
60 pF vac & Copper tubing	7.040	300,000	0
Coax RG-58	3.700	23,200	0
Coaxial with fixed mica capacitor	7.040	21,660	0

Highest R parallel equivalent is best!! Lower Rp means more loss.

Trap Measurement summary:

- **Coaxial trap poorest**
- **Once #10AWG wire is used, not much improvement**
- **Space-wound bare wire makes best inductor**
- **Transmitting-type capacitors noticeably better than capacitors made from coax**

10 Meter (Tribander) Traps

Type	Freq	R parallel	X
Coax RG-58	29.00	13,800	0
Mosley TA-33	30.64	43,100	0
Mosley Pro-57	27.46	66,080	0

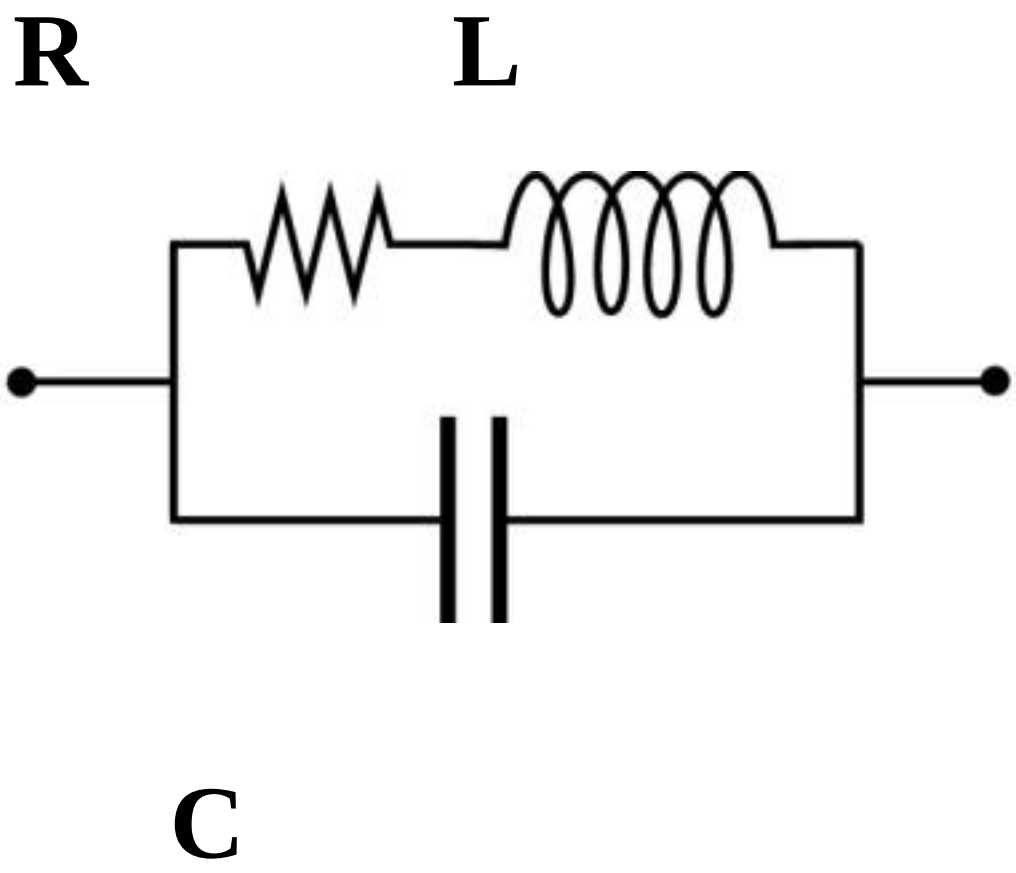
Cushcraft A3	**28.78**	**110,000**	**0**
HyGain TH-3	**29.67**	**140,200**	**0**

Traps are not all that bad when you plug them into models.

15 Meter (Tribander) Traps

Type	Freq	R parallel	X
Coax RG-58	**21.00**	**13,980**	**0**
Cushcraft A3	**21.43**	**76,270**	**0**
Mosley TA-33	**21.68**	**79,000**	**0**
HyGain TH-6	**22.23**	**142,000**	**0**

Trap Model



Measured Values Coax 7 MHz Trap

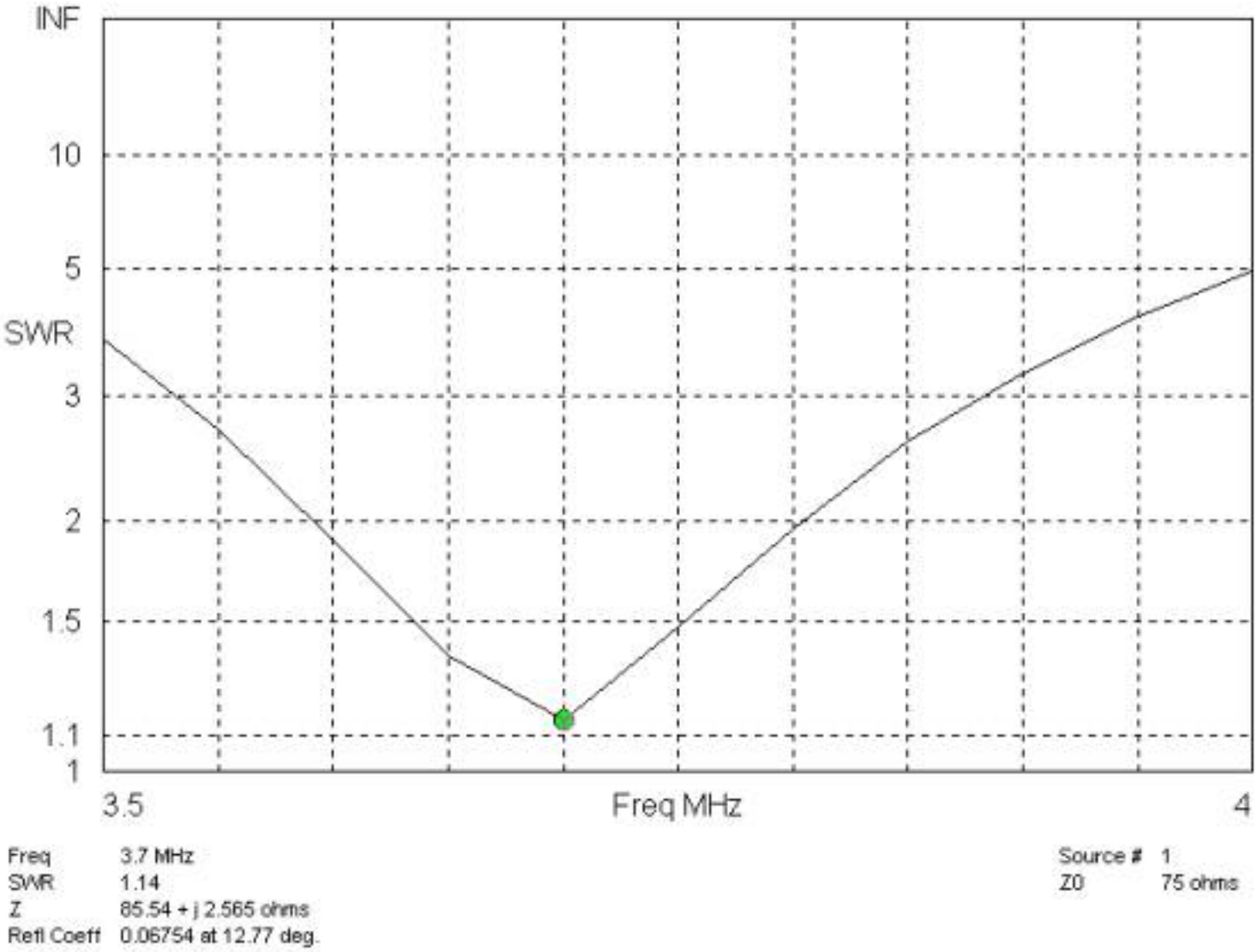
Freq	Imp	R	Xc	Q	L uH	C pF
7.04	17,800 j0	1.03	138	134	3.114	164

3.7	1.1 j 97	.6	283	88	3.114	152

Measured Values L/C 7 MHz Trap

Freq	Imp	R	Xc	Q	L uH	C pF
7.04	99,850 j0	.36	215	465	4.92	105
3.7	0.53 j 156	.25	409	294	4.92	105

SWR Bandwidth



7 MHz RG-58 TRAP

80 m 75 ohm VSWR

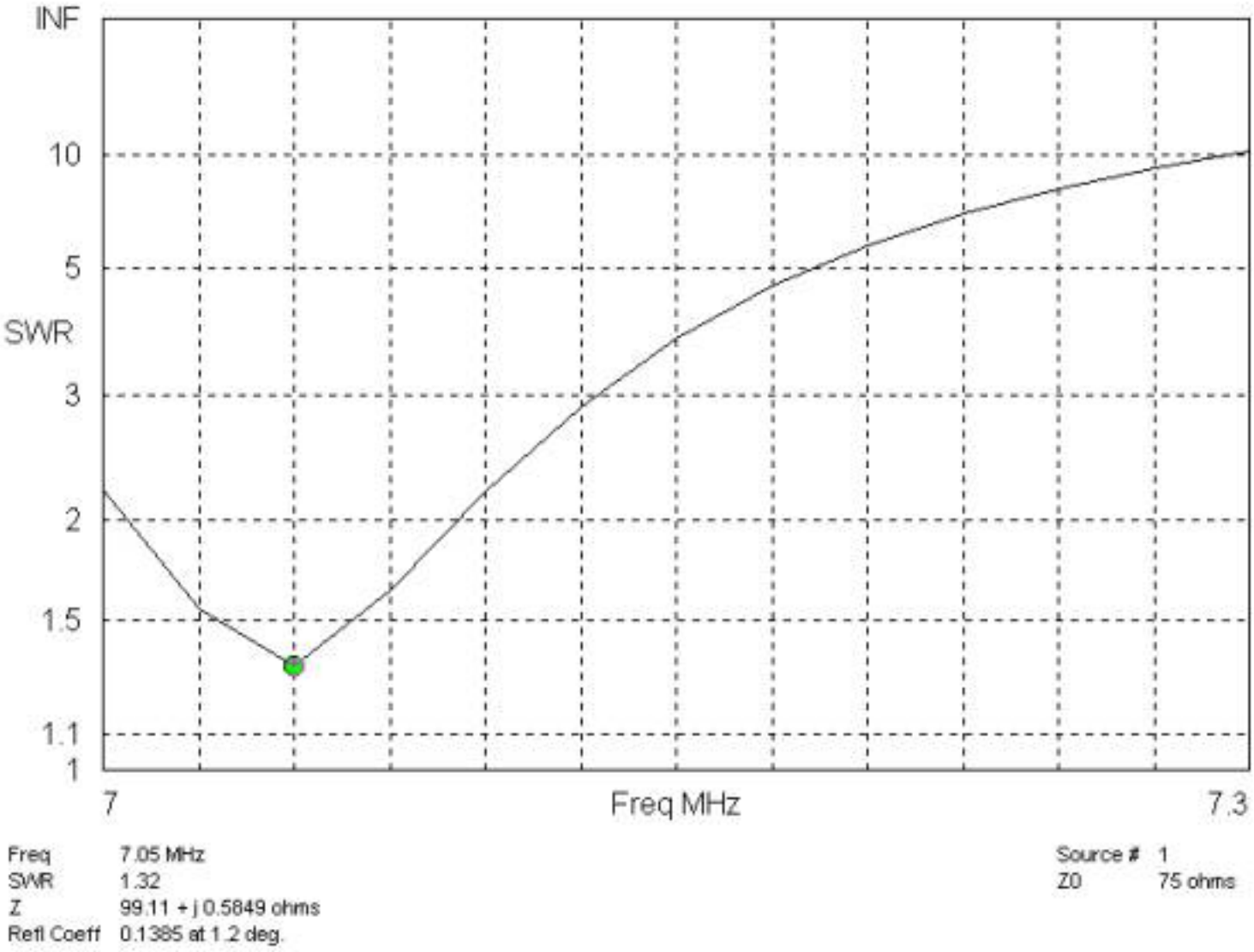
EZNEC #12AWG dipole

Coax trap 80m 2:1 VSWR ~210 kHz

Total trap loss = 0.05 dB

RG-58 TRAP, 75 ohm VSWR, 40 METERS

VSWR BW



Coax trap 40 meter 2:1 VSWR ~ 80 kHz

Total coaxial trap loss at resonance on 40m= 1.6 dB

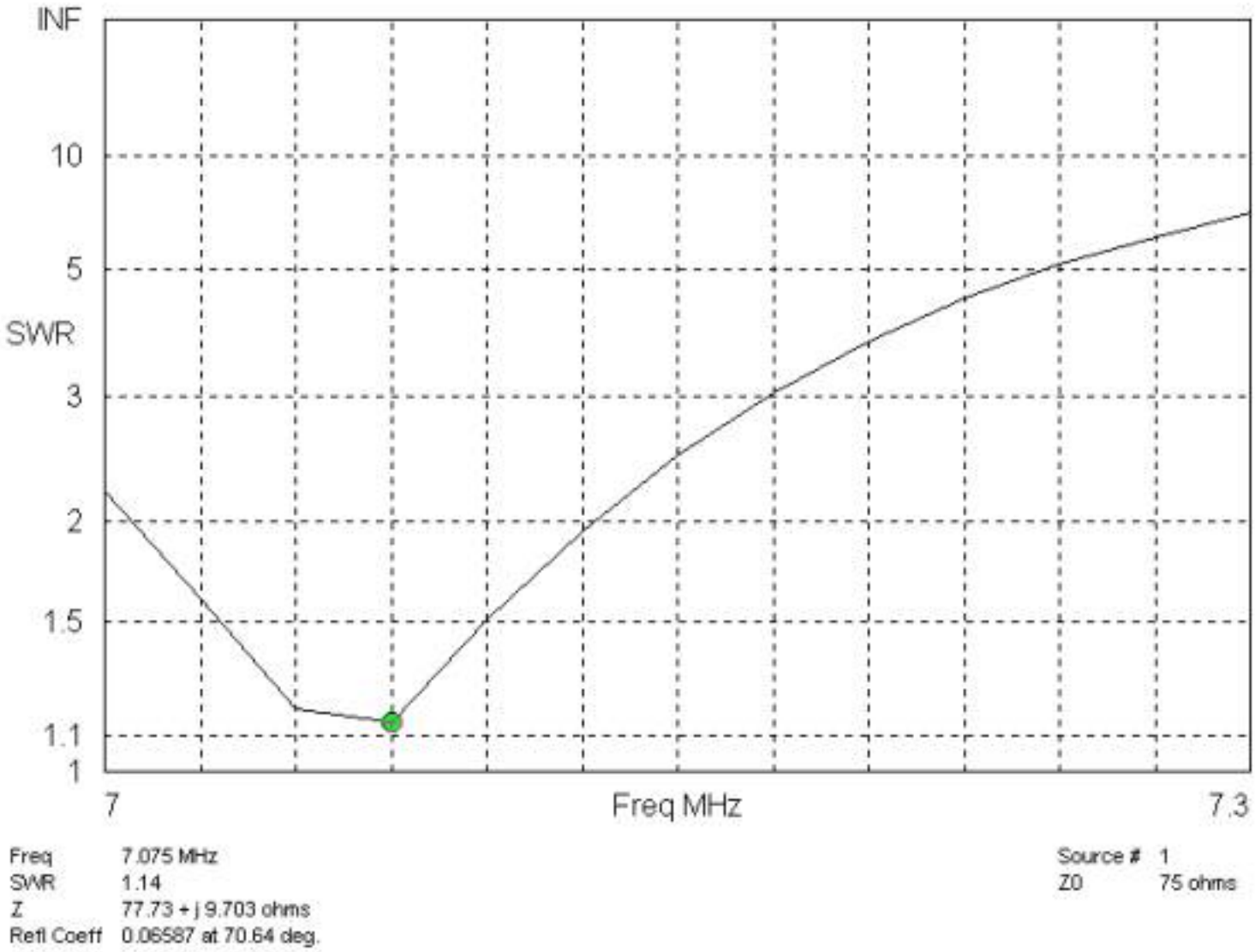
Total coaxial trap loss 100kHz off-resonance (at 7.15 MHz)= 1.06 dB

Note that loss is maximum at trap resonance!!!

Never make a trap resonant on the desired operating frequency!!

W2LH ARRL Handbook Trap Design

100pF #12awg Miniductor trap

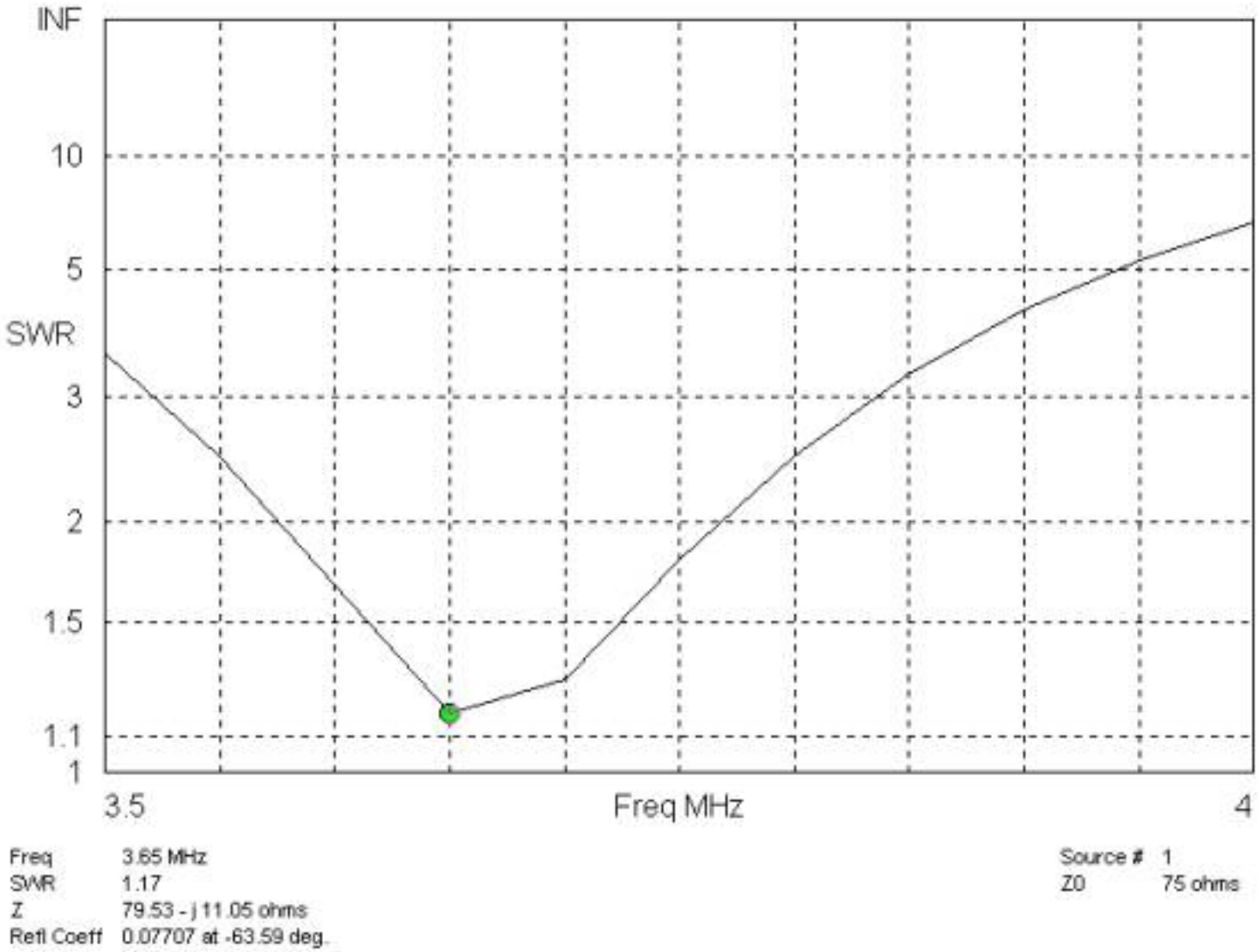


40m 2:1 VSWR ~120 kHz

Total loss = 0.24 dB

W2LH ARRL HANDBOOK TRAP 80m

VSWR

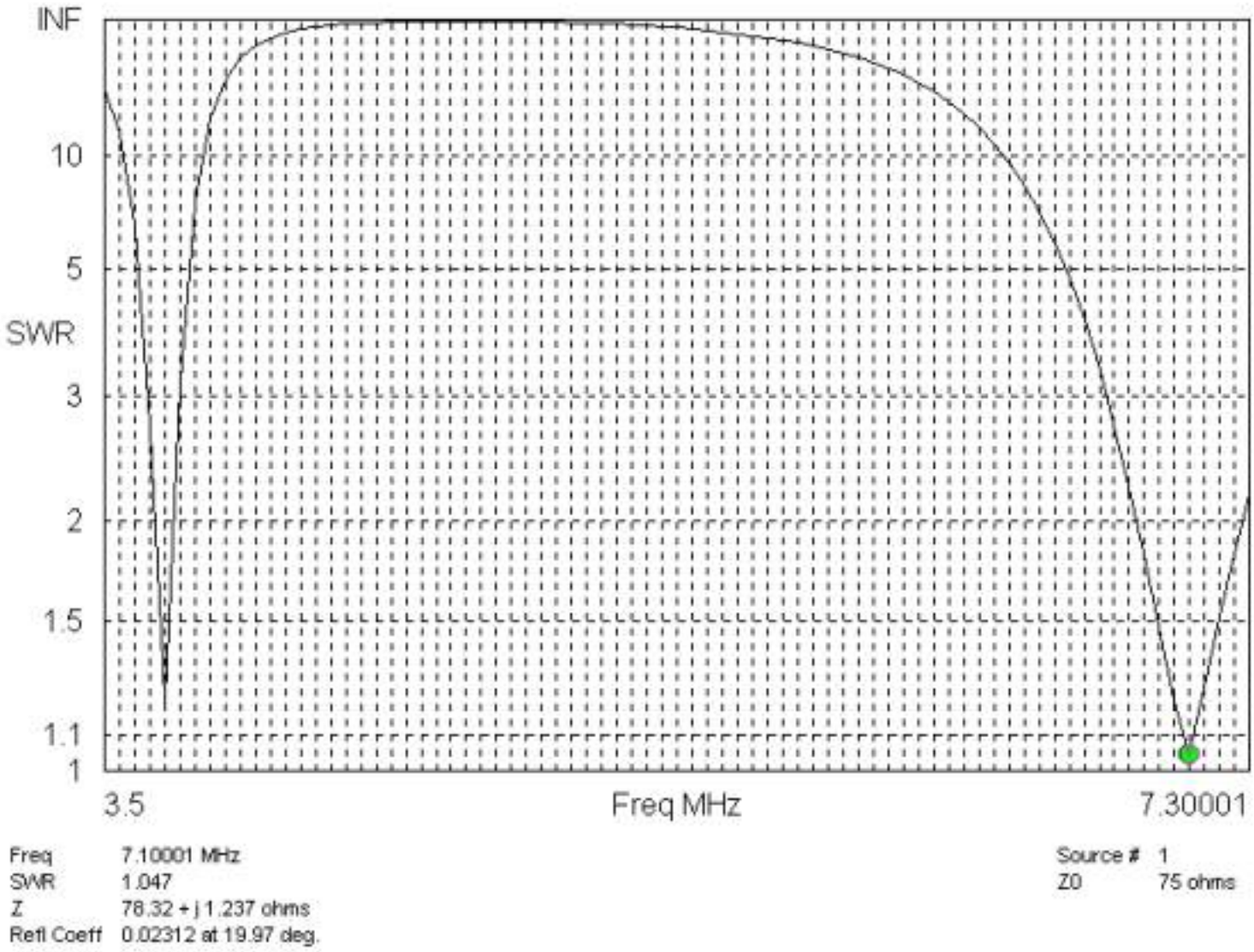


80m 2:1 VSWR BW ~ 200 kHz

Total trap loss = 0.026 dB

What happens if trap is not in band?

VSWR Bandwidth of 6.51MHz trap in 80/40 dipole



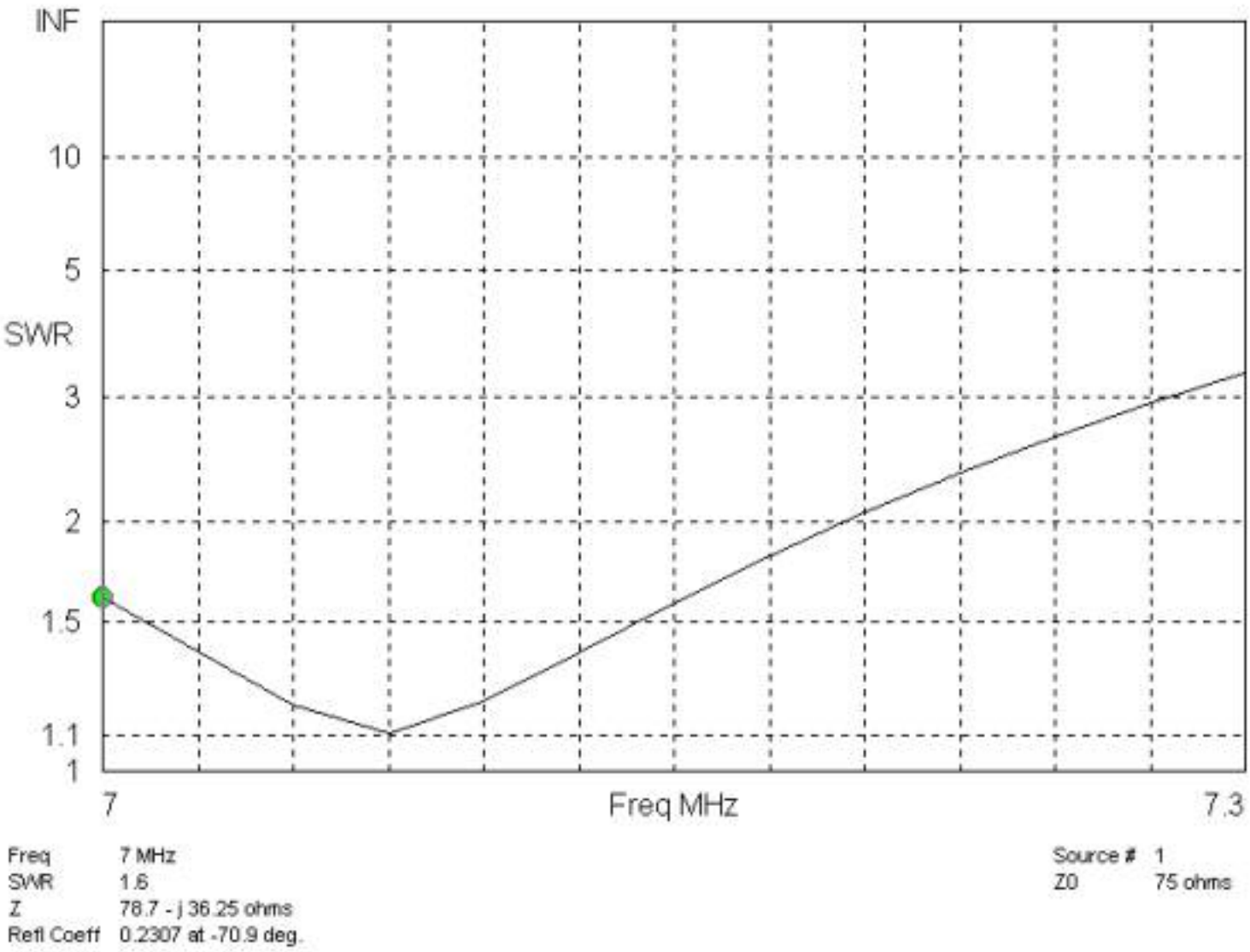
Trap at 6.51 MHz Q=130

Loss at 7.15 MHz = 0.314 dB

Loss at 3.7 MHz = 0.324 dB

This is a 104-foot long antenna, with very poor Q traps, and loss is less than .4dB! The reason loss is low is we have moved the trap slightly out-of-band.

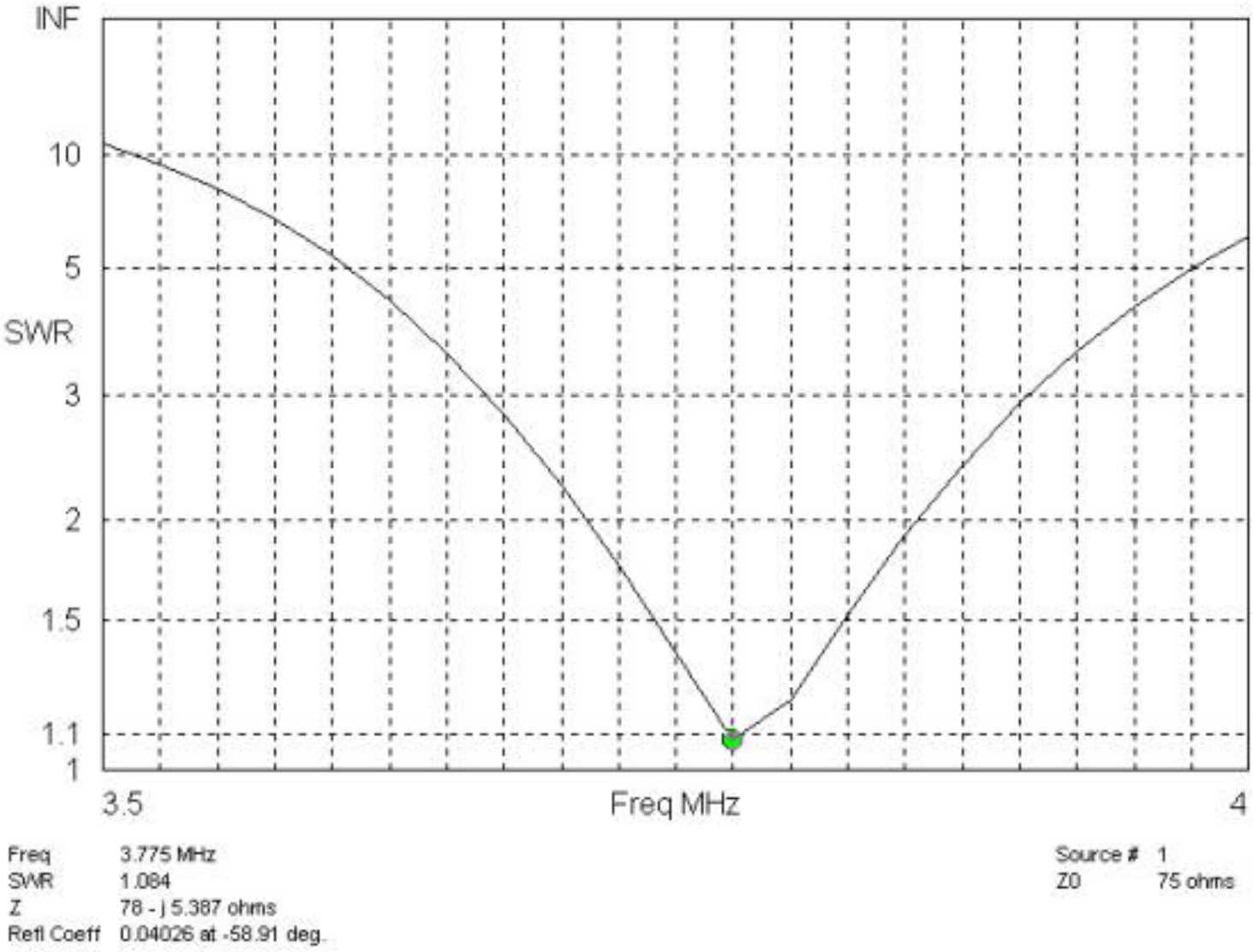
6.15 MHz Q=130 TRAP 40m VSWR



7 MHz 2:1 VSWR BANDWIDTH ~200kHz

Trap Q at resonance = 130 7 MHz loss ~ .3 dB

6.15 MHz Q=130 TRAP 80m VSWR



80M 2:1 VSWR BW ~130 kHz

Loss at 3.7 MHz = 0.324 dB

1.) Is it best to make the trap resonant close to the desired operating frequency?

NO! Loss is highest when the trap is resonant at the operating frequency!

2.) Does bandwidth decrease with increasing trap Q?

NO! Bandwidth is a function of many variables, trap Q actually has one of the smallest influences on BW.

3.) Do traps create noticeable loss, perhaps one dB per trap typically?

NO! Even the worse traps (coaxial traps) in the worse possible condition of operation are only 1.6dB loss for BOTH traps!

4.) Does higher trap operating Q always mean lower loss?

NO! Loss depends on many factors, including trap resonant frequency.

Conclusions:

- Trap loss has been greatly exaggerated by advertising hype
- Traps should not be resonant at the actual planned operating frequency
- Coaxial traps are more lossy than articles conclude
- Coaxial stubs used as capacitors can not be calculated using pF/ft unless the stub is a very small fraction of a wavelength long (less than a few electrical degrees)
- Coaxial stubs have low Q (are relatively lossy) compared to normal lumped components.

Building a Current Meter

[Home]

It takes a special meter to measure current accurately in the presence of strong electric fields, and to not disturb low-capacitance high impedance systems. The meter described below is suitable for measuring RF current accurately in high voltage applications, such as short antennas. It is also accurate in normal applications

Another current meter project link (***not*** suitable for mobile antenna measurements, however!)

http://www.ifwtech.co.uk/g3sek/clip-on/clip-on.htm

Low Capacitance, Voltage Immune, RF Current Meter

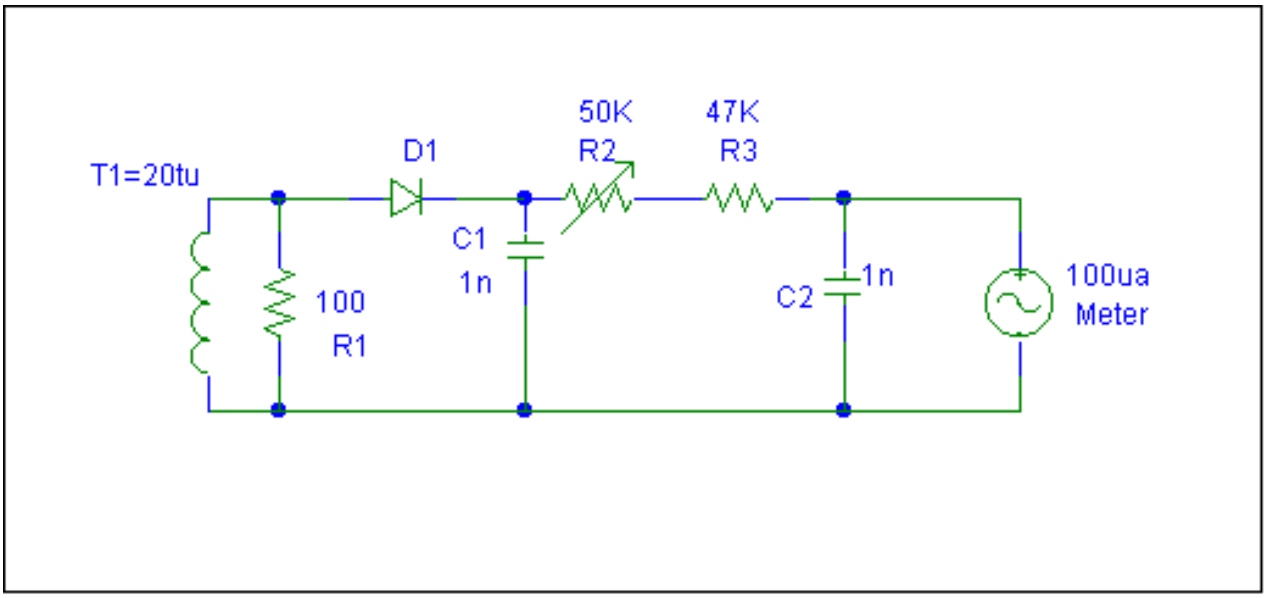
The meter is constructed with a 100uA all plastic case meter. The meter scale is plastic. The movement and all metallic areas are small. The lack of large metallic components minimizes stray capacitance added to the circuit under test by the proximity of the current meter.

Low capacitance ensures the meter has the least possible effect on the circuit being tested.

This meter has a 1.5" toroid current transformer on the rear. Just behind the meter is a calibration pot and all the circuitry. The toroid and circuitry is hot-melt glued to the meter with only the screwdriver calibration pot exposed.

I did not add a Faraday shield because the shield would increase the capacitance. Since this entire meter floats above ground, there is no need for a shield anyway, balance is not critical.

The meter has a low threshold Schottky diode detector, a series resistor for calibration, and standard .1uF bypass capacitors.



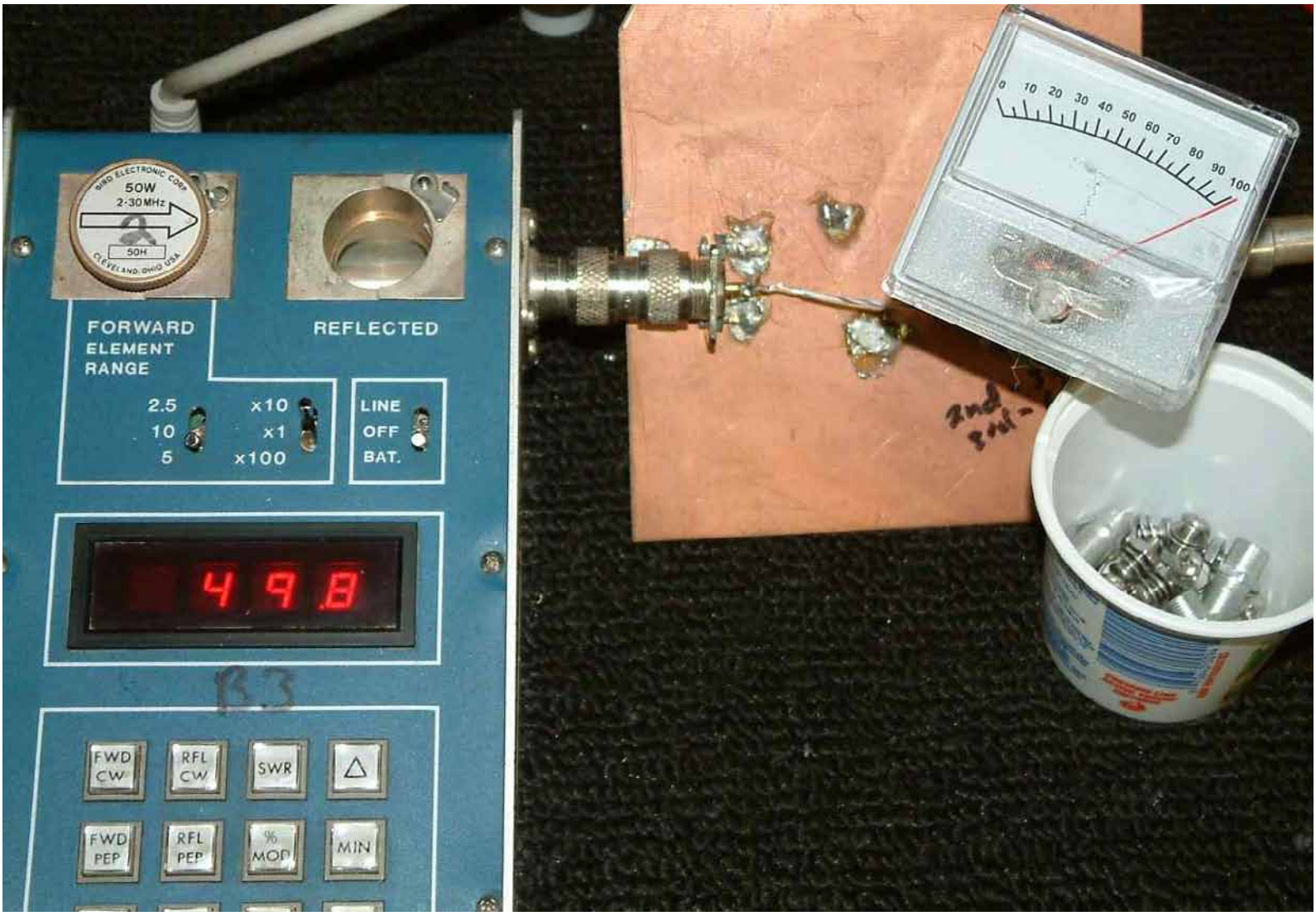
T1 is a current transformer. When the single turn primary (a whip or mast) has 1 ampere, the secondary will have

.05 amperes (inverse of the turns ratio). R1 flattens the response and limits the voltage. With 100 ohms we have .05*100=5 volts RMS. The peak dc voltage is 1.414 times 5= ~7 volts. C1 is a filter capacitor for the RF pulses, R2 and R3 set the FS range. With 100uA meter the resistance is 10,000 ohms/volt. 7 volts requires 70K ohms, which will be approximately midway on R2.

Dissipation in R1 is .25 watts from .05 amperes (T1 current) times 5-Vrms (secondary voltage of 5v with .05a flowing through 100 ohms).

Calibration

The meter is tracked in a test fixture for linearity. The actual fixture used is shown below.



The test fixture consists of two UHF female connectors soldered to a sheet of PC board. There is a single #16 Teflon wire running between the center pins.

Note the numbers "2" and "B3" on the meters. My elements and meters have calibration charts that correct for linearity errors in tracking and readings. In this case I've applied 50 watts to a precision 50 ohm load, making wire current 1-ampere.

The meter is calibrated and scale was tracked from 1 ampere down to .2 amperes.

FS accuracy is not required in comparison measurements, since the meter references against itself. Linearity within a few percent is important.

This type of meter is much more reliable and linear than thermocouple RF ammeters, and perturbs systems much less. Stray capacitance added to the system being tested is very small, because only the proximity of the meter and the compact wiring area. Contrast that to actually connecting a meter in the line with the associated lead lengths and capacitance of the meter itself, and the advantages of a transformer coupled meter become apparent.

Back to Mobile and loaded antennas.

RRAA_post

[Home]

I really begrudge the necessity of posting once more on this newsgroup, and particularly on this thread. However, I'm dismayed and disgusted by postings being made by Yuri and Cecil in other forums in which they're claiming that measurements I made agree with theories and predictions they allegedly made, and that my measurements therefore validate their theories. (A quick scan of this thread shows that they even made the false claims here, after I had quit posting.)

One of the postings is the following, made by Yuri on the eHam TowerTalk group, Nov. 21, topic "Trap Resonance":

"Why don't you mention what W5DXP came up as explanation for Tom's errors, why don't you mention what W7EL measured and that it was what I predicted based on available information and was right on - the 5% difference for the base loading coil?"

and this one, posted by Yuri on the EHam.net forum topic "Re: Current in Antenna Loading Coils" on Jan. 7:

"I will leave it here, as the rest of it. W7EL, author of EZNEC measured toroid coil and found that it HAS different current at its ends, roughly proportional to the part of antenna that it replaces."

and this one, posted by Cecil on that same group on Jan. 12:

"Roy's data clearly illustrates the phase shift through the coil. ARCCOS(Iout/Iin) gives an estimate of the phase shift (assuming forward current and reflected current are of equal magnitude). In Roy's experiment, Iout/Iin was about 0.95. ARCCOS(0.95) equals 18 degrees, an approximation for the phase shift through the coil."

As you'll see below (or by looking up the original thread), the first of my two measurements, for an antenna shortened an equivalent of about 18 degrees, resulted in 3% current attenuation across the coil (not 5%), and zero phase shift (not 18 degrees). The second test, where the

antenna was shortened more than 33 degrees, measured 5% current reduction and no phase shift. The method used in the above quote predicts more than 16% amplitude reduction and 33 degrees of phase shift for the second test. There's no way my data "clearly illustrates" Cecil's explanation. To say that it does is a pure fabrication.

I feel compelled to respond to these fabrications, and put the record straight. I'll do it here, since this is where my measurements were originally posted.

I made two sets of measurements of the current into and out of a toroidal inductor at the base of a vertical antenna. The details of the measurement method and the measurement results were posted here, on this newsgroup, on this thread. Pictures of the setup were posted on my web site, with a link posted here.

Before I posted each set of measurements, I asked for predictions of the results, so that alternate theories could be tested. (I was criticized for doing this -- it seems that the preferred method of testing a theory is to look at the results first, then adjust the theory to fit.) Yuri made a prediction (actually, two different ones) for the first set of measurements that didn't accurately predict the results. When I calculated the predicted result for the second set of measurements using the same method he had used for the first prediction, he retracted any claim that the method would be valid. (Exact quotes are below.) He didn't make any prediction at all for the second set of measurements.

Cecil made a number of vague predictions which he later contradicted or retracted. At the time the second set of measurements were posted, he had made no prediction at all.

Perhaps Yuri and Cecil have, after the measurements were posted, developed theories to explain the results. As of the time the measurements were posted, they hadn't. I highly recommend that anyone considering their alternative theories to find where they have calculated the results which agree with my measurements (particularly the second one, which was designed to produce a testable difference), and how they derived the equations used for the calculation.

Following is a summary of some of the exchanges between Yuri and me on this thread last November. The entire thread, "Re: Current in antenna loading coils controversy" and variants, is available for viewing at groups.google.com.

------ Summary ------

Here's what really happened. The following quotes are directly from the google archives of the rraa thread.

I made two sets of measurements. The first had the inductor connected at the base of a 33 foot vertical. But the vertical was mounted about 1/4" from a four foot pipe, which reduced the base reactance.

Here, I was asking for predictions for my FIRST measurement -- the one with the vertical mounted on the pipe.

Yuri posted on Nov. 9:

"In that case,
If the feedpoint current was at 0 deg of the radiator length, and coil replaces 18 deg of wire, the cos 18 deg = 0.951 which should make difference, drop in the coil current 5% (or half, 2.5 deg?)
Providing current maximum is exactly at the bottom end of the coil."

and later on Nov. 9:

">Incidentally, I take it that your prediction for the setup I did measure
>includes an 18 degree phase shift of current from input to output of the
>inductor?
>
>Roy Lewallen, W7EL
>

Yes, I used Cecil estimate/calculation and taking
cos 18 = 0.951056516 which is 4.8943483%"

So now we have his prediction, using the "cosine rule". The measurement I made showed about 3% current reduction from input to output, but with about 2% (the same amount within measurement error) also occurring when the antenna was replaced by a series resistor and capacitor -- that is, no antenna at all. So the 5% prediction was wrong. His prediction of 18 degrees phase shift, which wasn't present, was also wrong. When asked for the justification for the "cosine rule", he never offered any, so its origin remains obscure.

However, I saw that the value was too small to be convincing, which is why I devised the second test. The second test used a more ideal antenna, with more of the antenna being "replaced" by the inductor. The "cosine rule" would predict more than 16% reduction, and more than 33

degrees of phase shift.

Before I gave the results from the second measurement, I posted the predictions which had been made, as I understood them. Since Yuri had invoked the "cosine rule" for the first test, I naturally assumed it would also apply to the second. (This is simply applying the equation Yuri used in his Nov. 9 posting to the second antenna setup. It's also the equation now being used by Cecil, as shown in his quote from the eHam group.) So in my posting I said:

(Quote from my posting on Nov. 11):

"**Yuri's method predicts a reduction of output current magnitude of 16.5% and a phase shift of 33 degrees."

to which Yuri responded, also on Nov. 11:

"It is not my theory. My argument with W8JI and his followers: is the current in typical loading coil in quarter wave radiator same at both ends or does it drop with distance from the feedpoint. I have made temperature observations, W9UCW measured the difference, W5DXP provided some explanation. Based on Cecils analysis of data you provided, and on my understanding of the phenomena I guestimated drop in current in your setup. No theory, no mathematical procedure (yet) just attempt (using degrees replaced by coil in a radiator) at explanation of what is happening. I will measure things myself, try to verify previous measurements and then come up with conclusions and "theory". So far Cecils (and ON4UN book) theory seems to be closest to the truth. . . "

So now, Yuri has disclaimed the "cosine rule". /He made no other prediction of the results of the second test./

In summary, Yuri first stated that the "cosine rule" can be used to calculate the current drop. That would have predicted over 16% current reduction in the second test. Then he retracted his claim that that theory would work, before the results from the second test were posted, and never made any other prediction. He never predicted the 5% result which was measured, as he's now claiming.

And if you can find a numerical prediction anywhere in the thread which Cecil made and stayed with, my hat's off to you. $100 goes to the first person who can point to any prediction made by either Yuri or Cecil before the second measurement results were posted that predicted second measurement results of 5% magnitude and zero phase shift. (In the case

of Cecil, this would have to be a prediction that wasn't later modified or retracted before the second set of results were posted.)

My measurement results are consistent with the fact that the currents into and out of a physically small inductor are equal. The small magnitude difference I measured can be explained by stray capacitance on the order of 7 pF from the output to ground and/or the probe -- not an unreasonable amount to expect. In no way do my measurements support the odd theories being proposed by Cecil and Yuri, and any statement that they do is completely false.

Roy Lewallen, W7EL

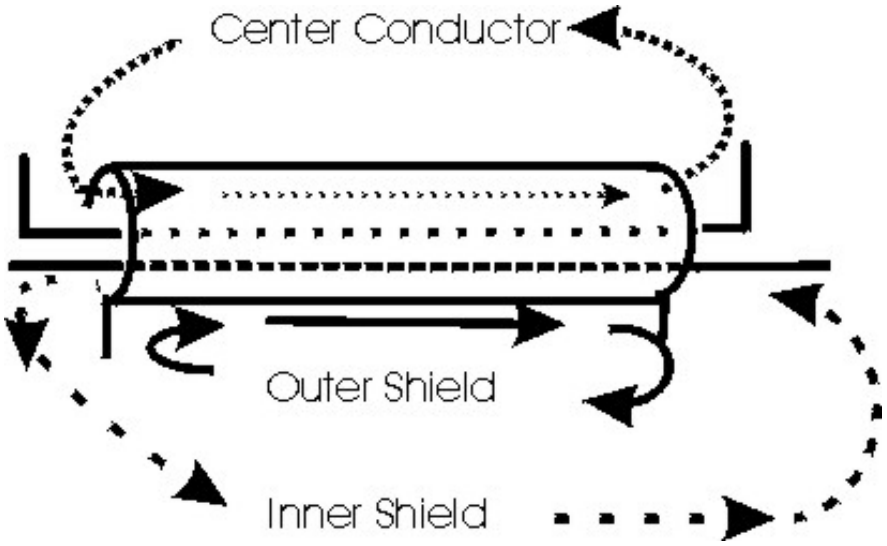
Coaxial line and shielded wires

[Home]

related pages at antennas

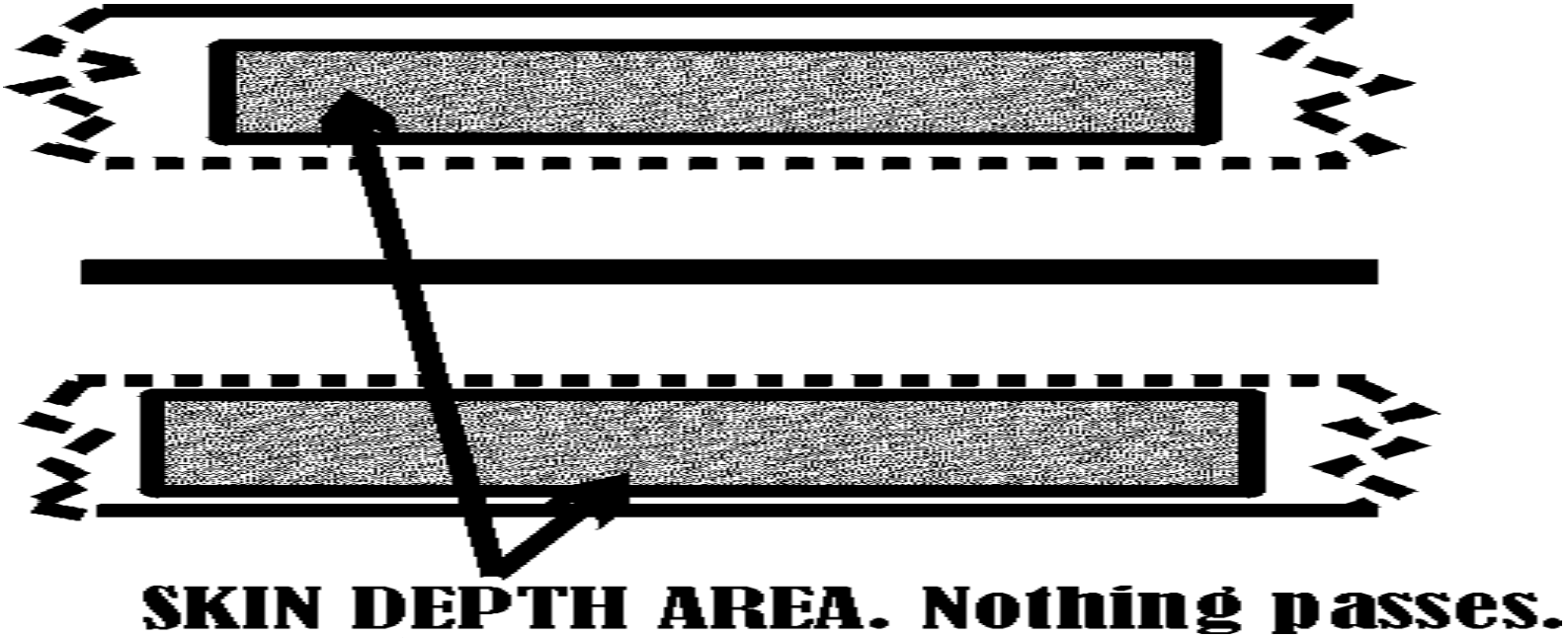
Concentric Transmission Lines (Coaxial Cable)

When the shield is more than several skin depths thick coaxial cables, single-wire shielded cables, or concentric lines become the electrical equivalent of three wire concentric transmission lines. Electrical rules require the center conductor and ***inside*** of the shield to always carry equal and opposite RF currents.



Notice in the drawing above the center and inside of the shield carry equal and opposite direction RF currents. This ALWAYS is the case when the shield is several skin depths thick. We cannot force anything else to happen!

In the drawing above and below, the outside of the shield is isolated by skin effect. It behaves like a separate transmission outer conductor. Skin effect prevents any current, voltage, or field (even magnetic) from penetrating the shield when the shield is many skin depths thick. Only the breaks at the ends connect the inner and outer shield conduction layers.



The only egress or ingress points for voltage and current are where discontinuity in the shield might occur through an intentional or accidental break in the shield.

Radiation In or Out of Shielded Lines

To operate without radiation, coaxial cables require equal and opposite currents in the shield and the center conductor. The outside of the shield must not have an RF potential difference along it's length. We cannot have voltage gradients (electrical fields) along the length, or the outer conductor layer will conduct currents. This means both ends of the cable should have zero volts to earth or the environment the cable operates in.

If we feed exactly equal and opposite current into the coax, currents have no choice but to flow ***INSIDE*** the cable shield! The outermost conductor can be treated as a single wire, since everything else is "hidden inside". There will be no "extra" current to flow over the outside of the shield. Because of this rule, we do not need to ground the shield to prevent radiation. The trick here is we need to have equal and opposite currents. When the shield has potential difference along the length we can greatly reduce outside currents by making the outside of the coax have a high impedance. We can do this by selecting a proper cable length, by adding sleeves of soft-iron magnetic materials or winding the cable in a coil with or without a core to form a choke. ***Skin depth prevents the inside of the cable from "seeing" what is done on the outside.***

Most interesting in all of this is the claim or idea the "shield" of a shielded loop allows magnetic fields to pass while filtering electric fields. The same rules apply as with coaxial lines. The shield actually becomes the antenna, and it couples to the inside only at shield gaps. If the gaps aren't exactly opposite the grounded part of the loop and inner conductors don't exit exactly at that grounded point, the shield actually UNbalances the loop!

Magnetic Loop Antennas Receiving

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(Please read the Radiation and Fields page)

Small Receiving Loops

Small loops are often referred to as "magnetic radiators". Folklore claims a small "shielded" loop antenna behaves like a sieve, sorting "good magnetic signals" from "bad electrical noise".

Nothing is further from the truth! At relatively small distances a small magnetic loop is more sensitive to electric fields than a small electric field probe.

The ratio of electric to magnetic field sensitivity is sometimes called the "field impedance". An antenna with ***high field impedance*** has a dominant electric field while an antenna with ***low field impedance*** has a dominant magnetic field when compared to free space field ratios at a large distance from the antenna.

Although fields have different ratios close to the antenna, at distances of ***about λ/2 the field impedances of all but very large antennas are virtually indistinguishable from each other***. Once we understand the basics, we might picture our antennas differently. We might find antennas don't respond to signals and noise like we thought, but at least we won't be as susceptible to the "noise" caused by rumors and folklore!

Loop Antenna Fields

It is the response in within λ/10 distance from the antenna that gives small "magnetic loop" and "electric dipole" antennas their names.

Very close to a small loop antenna (but not necessarily near the open ends of the small loop where the tuning capacitor is) the magnetic field dominates. Current is essentially uniform all around the loop circumference, while voltage has a nearly straight-line increase as we move to the tuning capacitor area. Since most of the area has uniform current and only a small concentrated area has the highest voltages, the magnetic field clearly dominates over most of the loop area.

Magnetic fields are effects derived from the actual moving of charges (current flow). The magnetic effect is related to current, and if it is proportionally large compared to voltage the field is described as having a "***low***" impedance. This is similar to the description used in circuits, where a system with high current and low voltage

is said to have "low impedance".

Small Dipole Fields

Near a small dipole or monopole the electric field dominates. A short antenna has to have very high voltages all along it's length. It normally has highest current only near the feedpoint, with current tapering down in a straight line to zero at the antennas open end. The dominant effect we deal with is from the very high voltages that appear all along the antenna. We say the antenna has a "***high field impedance"***.

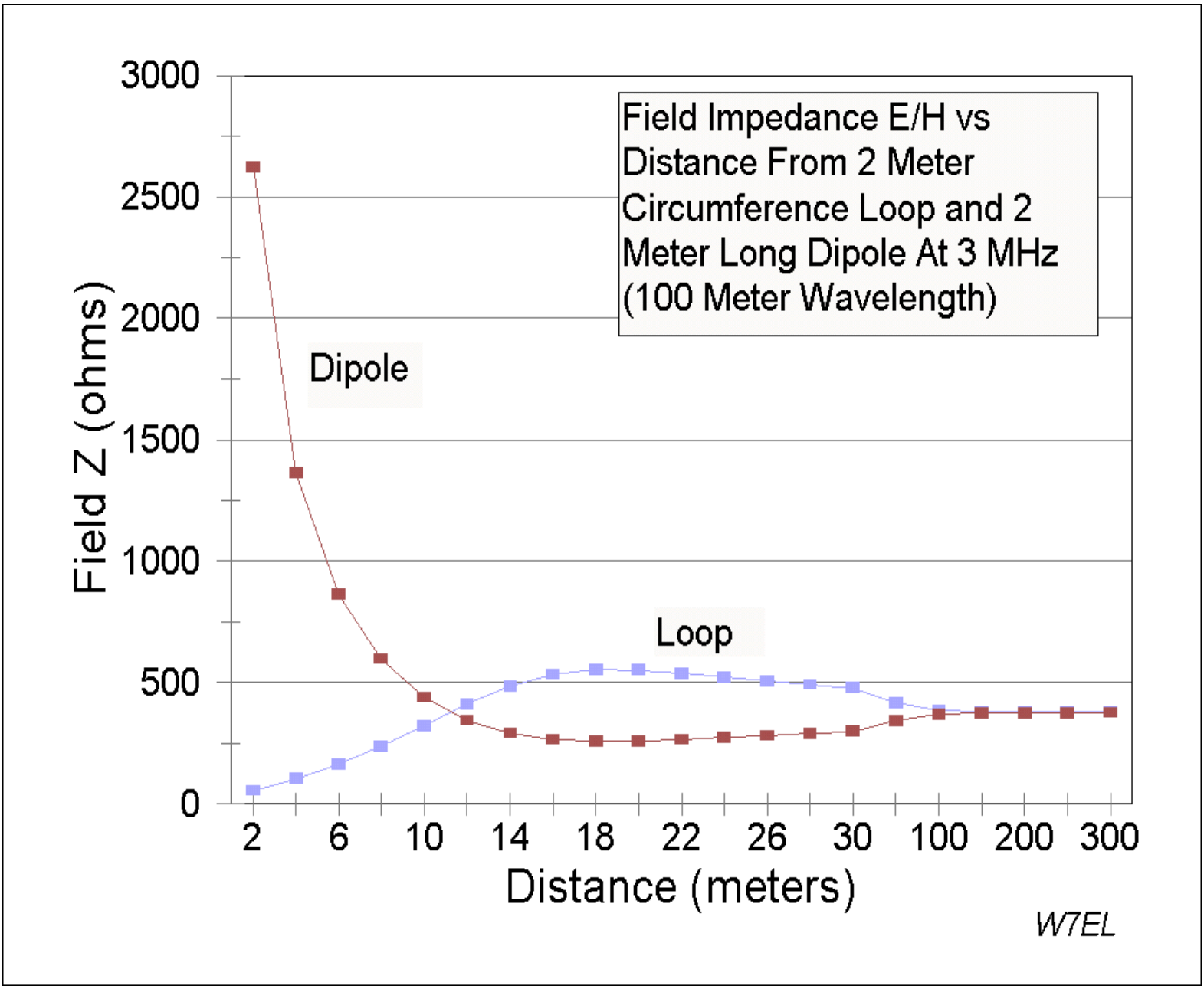
In the electrically small antennas described above, the dominant field descriptions only apply within λ/10 distance!

Radiation

The induction fields described above are NOT what radiates or receives at a distance. The induction effects are a necessary byproduct of:

1. Uneven charge distribution (a difference in voltage) causing a physical force on other charges. We call this effect an **electric field.**
2. Moving charges (a current) causing a force on other charges. We call this effect a **magnetic field.**

Accelerating charges causing a force on other charges. We call this effect **electromagnetic radiation. It is a totally different effect independent of the induction fields. This is the only effect that works at a distance.**



Since the distance of a wavelength in the above graph is 100 meters, we can considered the bottom scale as the percentage of a wavelength. We can see at about 11 percent of a wavelength (which would be about 50 feet on 160 meters), there is no difference in field impedance between a small loop and a small dipole. At distances beyond 50 meters on 160 meters, the magnetic loop actually has a ***higher*** field impedance than a dipole.

So much for the myth that a receiving antenna can sort good signals from bad signals (noise) by virtue of being "magnetic"! We not only don't have the response we might have imagined, we also almost certainly have no idea if close-by unwanted noise or signal sources are radiated from electric or magnetic field dominant sources. Successful noise reduction by virtue of by antenna "style" would mostly be a matter of hitting a lucky combination through careful experimentation.

Loop Shielding and Balance

Loop shields do not sort noise out, nor do they prevent electric fields from affecting the antenna. For a description of how shields work, look at the Concentric and Coaxial Transmission lines page. From that page, you will see the shield actually becomes the antenna in a "shielded" loop.

There are many construction articles about small loop antenna available. I'm not going to add one more to the group, but I can tell you what to watch for. It is *VERY* important that all conductors exit the loop at the ground point of the shield, and that the loop is grounded exactly at the electrical center of the shield. The loop must also be symmetrical, and you must mount the loop so the feedline and any metallic supports leave the center area of the loop with maximum symmetry. If you DON'T do this, the loop can actually use the feedline as the antenna. This can greatly increase sensitivity of the loop to conducted noises!

Remember the following guidelines:

- The shield is the actual antenna
- The shield must be perfectly symmetrical away from the inner wire exit point
- The gap in the shield must be exactly opposite the grounded point
- The ground must be at the inner wire exit point
- The shield will not make an unshielded loop that is properly balanced any quieter
- ***The shield only is a tool to help you balance the system IF the shield is properly implemented***

I hope this article is useful in helping you select a GOOD design!

E-H Antenna

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Background

Many antennas function because of common mode currents, rather than working in spite of them. Two popular examples are CFA and EH antennas. Another recent example, appearing in Antennex's compact antenna articles, is a thick stub "vertical" with no counterpoise.

All of these antennas become significantly poorer radiators if common-mode currents on feedlines are eliminated. Why? Because the feedline is the actual radiator, ***NOT*** *the tiny thing they call the antenna.*

Misunderstanding or misapplying Maxwell's equations and the principles behind radiation, in combination with missing some very key points of conventional circuit theory, causes problems. Some of us have unwittingly attributed increased feedpoint resistance and/or seemingly disproportionate amounts of radiation from very small structures to new methods of radiating EM waves. Reviewing these antennas and the theoretical or technical mistakes surrounding them will help us understand how antennas and transmission lines work. With that knowledge, we can build better antenna systems. The fastest and best way to learn is often to look in detail at mistakes!

What They Claim

Articles and user reports of CFA, EH, and thick stub verticals (without groundplanes) appearing in Antennex and other internet publications have one common thread, the operational descriptions almost always include strong indicators of problems with feedline common-mode current.

Authors commonly warn users to NEVER choke feedlines with baluns and to "be sure the feedline is straight and in the clear"! Authors lay blame for RF burns from feedlines or shack equipment on the antenna's "high radiation efficiency", claiming these small magical antennas radiate so efficiently they naturally excite the feedline and equipment more than full-size antennas.

For example, Ted Hart (of EH Antenna fame) claims the following at:

http://eh-antenna.com/documents/RF_on_the_Coax.pdf

"RF on the Coax

Due to the large radiation at the EH Antenna, there will be some RF coupling to the coax. Whether this is a problem is dependent on the radio you use. Some are subject to RF coupling into the audio system, which causes severe distortion while transmitting. On some field day setups with 100 watt transmitters we have had so much RF on the radio you can get an RF burn. Below we have suggested ways to eliminate the RF coupling problem."

In the above statement, Ted actually acknowledges current on the feedline shield and RF-voltage-on-chassis problems. The problem must be severe when a low-power 100-watt radio causes a burn. Like any good salesman, he turns a design shortfall into a feature! According to the author, unwanted RF on the feedline doesn't come from a feedpoint or antenna design problem like it does on other antennas, in this case it appears because the antenna works so well!

Here's what actually causes RF to appear on a coax shield and radio chassis. RF can only appear on the radio chassis through two methods:

1.) The antenna, from poor feedline or feedpoint design, can couple to the radio chassis through external wiring or cables attached to the radio.

2.) The radio chassis itself, being large in terms of the wavelength, can actually become an antenna and receive energy from actual desired "over the air" signals.

(Many of us have these problems. Click on this link to see one reason why.)

In this case, we can probably rule out reason two above. It is unlikely the chassis is a large portion of a wavelength long on 20-meters and that the antenna field is suddenly so strong it is "lighting up the house" with RF. After all, if power is radiating effectively it will all be going out to distant stations and NOT cooking you or the radio gear in your house! That only leaves reason one, poor feedpoint and antenna design, as the cause of common-mode feedline or wiring currents that excite the coax shield and eventually the radio chassis, as described in method one.

Actually someone has measured this, and posted it to a web page!

http://www.home.earthlink.net/~calvinf15/_technical/

When the time-varying current from the transmitter flows in any conductor, we will have charges accelerating in the cable. The outside of a shield is no exception. A feedline's shield will radiate proportionally by the ampere-feet of the cable. just as any other conductor will. Of course the antenna element will radiate also, but there is something else to consider.

A very small current flowing unopposed over a large linear distance will radiate quite a strong signal, because radiation resistance of a long antenna is generally very high compared to very short antennas. You can find this explained in the Radiation&Fields and Radiation Resistance articles on this site, and in engineering textbooks such as those written by Jasik, Kraus, and Jordan-Balmain. From all of this, we know the shield radiates.

The inventor of the E-H antenna goes on to say:

"If you use RF beads, since the coax shield is not a magnetic shield, the beads affect both the inner and outer conductors. Therefore, most of the transmitter power will be converted to heat. Not good."

Not a very knowledgeable statement at all, at least from the standpoint of how shields work!

Time-varying fields can not pass through a shield that is more than several skin depths thick. The inner part of the shield and outer part are isolated by the skin effect, and nothing passes through. The ARRL Handbook, Maxwell's book "Reflections", Reference Data for Radio Engineers, and dozens of other amateur radio and engineering texts describe this effect correctly. If we bring a time-varying electric field to zero in a system, the time-varying magnetic field is also by definition zero. The shield DOES isolate the center of the cable from time-varying magnetic fields on the outside of the cable!

If that is true, why then does a shielded cable passed through a current transformer used in a directional coupler appear to pass RF magnetic fields? Why does the RF magnetic field seem to "pass through" the shield of a shielded receiving loop antenna? The answer is quite simple. There is a gap or intentional break in the shield.

Current on the inside of the shield "spills over" the edge of the shield where the shield is broken, and causes a current on the outside of the shield. There is also a voltage across the gap at each end of the shield. We have both our time-varying voltage and current, via a circuitous path to the shield ends! Our Amateur texts explain that effect, as do all of the engineering texts dealing with shields on transmission lines. If the gap is closed and the shield's ends are shorted together, making voltage across the shield gap zero, the magnetic field no longer "seems to" penetrate the shield!

If a shield did not behave this way, we'd be in serious trouble in the radio world. Without the shield stopping both magnetic and electric time-varying fields, we could not shield our radios. We also could not shield our microwave ovens, with non-magnetic materials!

If anyone thinks a ferrite core affects the impedance inside a coaxial cable that does NOT have significant common mode current, they only need to slip some beads over a cable on a working normal "cold-for-RF" feedline. You will find absolutely no difference in system performance when

beads are added, proving Ted does not "have it right" in the text above that appeared on the E-H antenna web site.

The author and inventor of the E-H goes on to say:

"Use of a small choke made of several turns of the coax is good. We find that a wire connected to the ground side of the coax at the antenna and connected to either a ground rod or a wire laying on the ground will eliminate RF problems - in most cases. For some radios we also need to add a ground wire to the radio."

Of course adding a ground wire might help! The ground wire becomes the path for common-mode currents, or at least a portion of them. The additional wire to ground becomes part of a long-wire antenna (made by the cable shield) that actually does most of the radiating!

"A preferred method is to run the coax to ground then back to the radio. Near ground, connect the shield of the coax to a ground rod or radial. Another method is to connect a wire from the radio to ground. If the radio is very far from ground you will need to add a series resonant circuit in the ground wire to effectively cancel the inductive reactance."

It is understandable why this is a preferred method! The outside of the coax shield can remain the primary antenna, saving us the bother of installing the additional ground wire that becomes the antenna in the previous suggestion!

"It may take one or more of the above to solve your problem. Remember that if you have a good ground on the antenna, you have also minimized problems with lightning."

In the above text, we can see every solution carefully avoids installing an effective choke balun on the feedline. A properly designed and effective choke balun has no effect on a coaxial feeder or system SWR, unless the feedline is radiating!

Let's look at another E-H antenna experimenter and former proponent of the E-H antennas test of a 160-meter E-H antenna. You can read the text directly at:

http://www.qsl.net/iz7ath/web/02_brew/18_eh/english/pag11_eng.htm

Steve writes:

"With my short (and easy) tests I deduce the EH lose about 10 dB versus my short 10 mt vertical with capacitive hat in the top;
10 db is a big amount of power, but remember we are using 2 small cylinder of 37,5 cm on 160 mt band; my vertical performed better than a 40 mt long sloper, so, may be, loss versus full size dipole is not too much."

Steve claims the E-H antenna looses about 10dB compared to a 1/4wl sloper, which has an unpredictable efficiency. A typical sloper 1/4wl sloper is likely only around 50% efficient in the best situations, and more likely much worse! The actual range can be from 10 to nearly 100% efficiency. Steve's data repudiates Ted's claim and the CFA inventors claim that "crossed-field" antennas provide high efficiency.

"There seems to be something good;
May be optimizing something we can have more gain (or, better, we can lose less dB);
This antenna can be useful on the low band, expecially on 137 KHz (the EH cylinder should be 6 mt high with a 2mt diameter).
On the web I've seen an other similar antenna (ISOTRON), wich seems caming (as EH Antenna) from CFA antennas;
On the italian Radio Rivista (1995-1996) there were same informations about an "Antenna Toroidale" wich in same way remember the EH: will be interesting build and test one."

Steve seems to be saying something we all agree with. A short antenna will radiate, but not nearly as well as a full-size antenna.

"By the way, I think that my EH is not for Dxing; it will be useful for that radioamateur who have no possibility to install "long" antennas for low band, but want just to have local QSO or few contest-contact."

I agree! 10-20dB loss from a full size antenna would not make a good DX antenna! Now here come the current-on-the-feedline problems:

"Other point: tuning;
I think it's not as simple as others say; I've used an MFJ259B which tells me all about the antenna, but the tuning was critical. On the other side construction is easy;
I want suggest to you , when tuning, to connect the MFJ259B to the shack ground (if there's one). If not, EH will probably resonate higher (20-30 KHz)."

If grounding and ungrounding an antenna analyzer or any other piece of equipment connected to a coaxial cable causes resonant frequency or SWR of an antenna to change, the system has severe feedline radiation problems. See my article on testing baluns.

"Coaxial cable influence: inserting more coax cable, the resonance seems to vary a little; I suppose that's normal;

I don't know for certain if the line radiates or not; I've added two iron-powder coil forms back my TX and EH performance was unchanged. Then later I've added 2 big ferrite beads with a lot of turns on the roof (at the feed point of the EH) and all has changed. My signal was 3 S point weaker than before and S.W.R was very high.
I don't know if that loss is due to the changed resonance in EH (as I said before, outside the bandwidth performance goes down quickly)."

We see again that any attempt at reducing feedline radiation results in an antenna that does not radiate very well. With the feedline choked, Steve's EH antenna dropped 3 S-units in addition to the original 10dB from his sloper, or maybe 25dB or more!

There surely is a hidden message in all of the above contradictions!

How the E-H Antenna *Really* Works

There are many examples where designers intentionally use common-mode currents. Examples are found in textbooks, such as the "Antenna Engineering Handbook" by Jasik on and around page 22-6.

The antennas at the right, copied from Jasik's textbook, outline the derivation of a skirt collinear antenna from a simple feedline with the open end terminated into a conductor. (It could be a ground

Two or even three skirts may be added to the mast as shown in Fig. 22-5c, but reduced excitation of the lower sections diminishes the effectiveness of each additic skirt.[6] Thus, while the multiple-skirt coaxial antenna resembles a collinear arra in-phase half-wave elements, its gain is not as great. Typical gain values are ta lated in Table 22-2, but since no test data are available for this type of antenna, values given are merely engineering estimates.

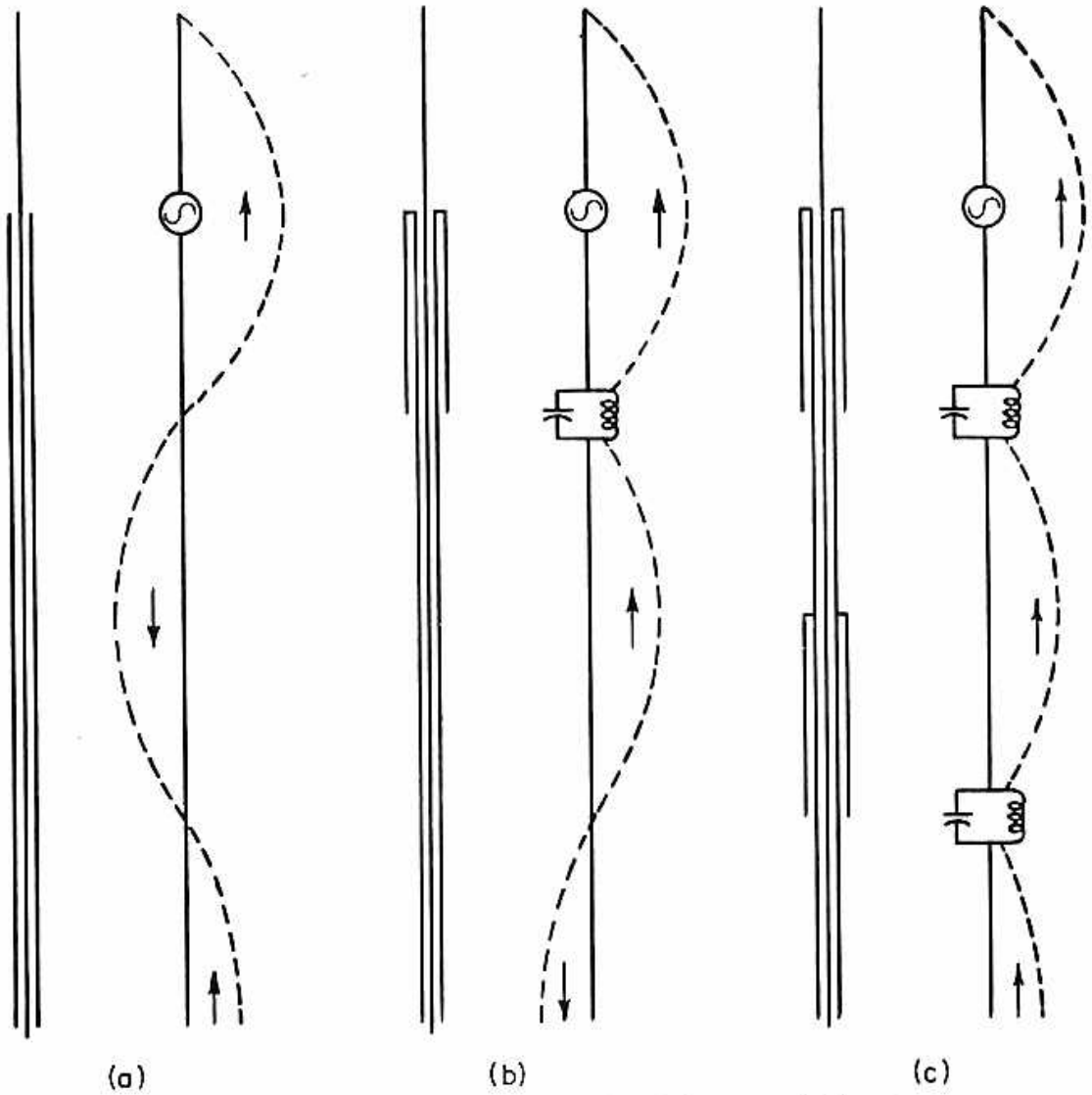

FIG. 22-5. Evolution of multiple-skirt coaxial antenna.

The relatively large diameter-to-length ratio of a practical skirt produces an effect which requires that the exterior length of the skirt be reduced by a factor of 0.8 to 0.9 and the interior length by from 0.95 to 0.98.[6] This is accomplishe cutting the skirt to the proper exterior length and inserting a dielectric slug ir skirt to increase its electrical length.

rod or an antenna, like a Beverage or large loop, the antenna does not have to be an "open circuit".

Looking at (a), we find by hanging any conductor from the end of a coaxial cable the shield is excited (on the outside) with common-mode current. The electrical equivalent of the OUTSIDE of the shield is just as if a generator located at the end of the shield was driving the outside of the shield as longwire antenna. This goes along with Kirchoff's Laws, that tell us the sum of currents entering a point must equal the sum of currents leaving that point. For any current to flow up into the antenna, and equal current must flow back down over the outside of the shield.

With one ampere flowing up the center conductor into the "stinger" at the coax's end-point, the same level of current flows back over the outside of the shield. (The shield's inside and outside are isolated by the skin-depth of the current at the operating frequency, an can be treated as two independent conductors that are connected over the open edge of the shield.) We *MUST* have this current simply because this is how coaxial cables work, the current on the inside of the shield is *ALWAYS* equal and opposite to current in the center conductor. There has to be some place for that shield current to flow, so it makes the bend over the end of the cable and flows back down the outside.

This is also why, when we use a cable's shield as a ground lead the center conductor and inside of the shield do nothing to reduce resistance. Any current that flows down the center conductor is cancelled by current flowing on the inner wall of the shield, the result is no current at all flows down the center conductor as long as the shield is several skin depths thick.

Many antennas intentionally and unintentionally use this principle, two examples are shown in (b) and (c).

A recent Antennex Article on a "magical" ultra-compact antennas claims an identical system, using a loaded fat cylinder, has an extremely high radiation resistance and excellent performance because some magical field-trickery increases the radiation resistance of a thick cylinder at the antenna end. Certainly radiation resistance is somewhat high...but not for the reason something magical or special is happening!

The small coil-loaded cylinder is actually only a fraction of the antenna length, and being so short has a very small radiation resistance. The point missed is the shield of the cable is very long, and is in series with that short section. Since the shield is long, it also has a reasonably high impedance both from radiation resistance and loss resistances. Shortening the length of the end-stub results in an insignificant reduction in radiation resistance, because the overall length of the radiating system is very long! We have a simple off-center fed dipole, with one very long leg and one very short leg!

The main radiator is the outside of the feedline shield, not the tiny thing being called an antenna!

Unless we make the coaxial shield an infinite length or pass it through what amounts to an infinite groundplane with zero resistance, current continues on down the cable shield. Looking at (c), we find even multiple sleeves appearing as parallel tuned high-impedance circuits do not fully decouple the

shield. Many collinear antennas work on this principle, yet E-H antennas and others attribute it to some form of electro-voodoo.

Moved Here

Ground resistance measurements

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160-meter Ground Connection Resistance at W8JI

Clay soil, dry conditions, 20-foot tall hat element resonant on 160 meters. All radials 50 ft long except for test radial, which is 1/4 wl electrical insulated counterpoise.

Ground	Ground Resistance Ω	Antenna Resonance MHz	System change with resonant test radial	System change adding one additional rod
Two 5’ rods	85	1.944	Large over 50%	Rod has large effect
One radial NE	80	1.905	Large	Rod has large effect
One radial SE	80	1.905	Large	Rod has large effect
One radial NW	77	1.931	Large	Rod has large effect
Four radials	42	1.819	Moderate 50%	Rod has large effect
Three radials and two rods	50	1.819	Moderate 50%	Rod has moderate effect
Four radials three rods	54	1.820	Low 25%	Rod has noticeable but small effect

Seven radials three rods	29	1.875	Very low <10%	Rod makes very little difference
Eight radials three rods	27	1.880	Very low	Rod makes very little difference

Performance Indications of Test

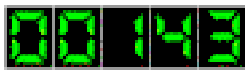
Ground systems are reactive. Poor grounds can add reactance. This is shown by the resonant frequency shift as the ground system is altered.

In slightly moist clay pastureland soil it takes about 8 short (1/8th wl) radials to stabilize the ground impedance.

In slightly moist clay pastureland soil, two or three rods make a poor RF ground for low impedance antennas.

If you use four short radials on a vertical, it is almost like no ground at all.

With small radial systems, ground rods can actually increase feed resistance. Use this particular phenomena with some caution, since I did not sweep frequency!



Echelon-Log Beverages

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Enhanced F/B Beverages

Two Beverages are paralleled a short distance apart, and stepped one in front of the other, can be phased to remove unwanted rearward response. This really just amounts to end-fire phasing two antennas. It is effective with long antennas, and also increases F/B ratio (and directivity) with very short beverages (even those approaching 1/2 wl).

Many or most previously published systems use frequency-selective phasing systems. With such systems, operation on more than one band requires switching systems and many components. "Conventional thinking" will make the system more complex than needed, and actually provide less performance! This is unfortunate, because it is possible to design a simple phasing system with almost no components that will cover octaves of bandwidth...no switching required!!

The two basic schemes I've used are very simple. One system is totally independent of grounds at either antenna end, the other system requires grounds but is more forgiving of dimensional errors in construction. I've described a model of a loaded Beverage, for those who would like to experiment with a small-space Beverage at Slinky and Loaded Beverages.

Bandwidth of Directivity

In high-efficiency unidirectional ***transmitting*** phased arrays, element impedances change significantly with frequency and phase angle changes. In unidirectional high-efficiency transmitting arrays, mutual coupling causes each element in every end-fire element group to have greatly different feedpoint impedances. This seriously complicates the design of ***proper phasing and current distribution*** as frequency is varied. It is very difficult to cover a few percent of the operating frequency, let alone multiple bands, with a high efficiency array.

Most amateur receiving antennas copy what we do with transmitting, and many of our transmitting systems copy narrow-band single-frequency systems. This is very unfortunate!

The key to broadband performance is having frequency-independent stable element impedances, and similar element impedances, throughout the system. This requires thinking "outside the box", not copying of transmitting phasing systems (often TX systems that are not that well-thought-out or planned).

Requirements for Broadband Directivity

There are three ways to obtain stable predictable impedances or broadband directive performance:

- Multiple elements can be used, such as those in log periodic or Fishbone arrays
- Active elements that are far from resonance can be used (short ***non***-resonant verticals or small ***non***-resonant loops with amplifiers)
- Very lossy elements (even large elements) can be used, with losses swamping out mutual coupling effects

While all three methods work, the simplest systems generally use lossy elements that are large enough to provide sufficient external signal levels to overcome receiver system hardware noise.

Broadband phasing systems are easily implemented in systems where feedpoint impedance is stabilized through intentional or "natural" loss mechanisms. Losses are made large enough to "swamp-out" or dilute mutual coupling and resonance effects, antenna feedpoint impedance remains stable and predictable even when elements are end-fire phased with a unidirectional pattern and close spacing!

Beverages As Elements

In the case of Beverages, ***radiation resistance*** is very low (in the order of an ohm or two). Most of the resistance we see at the feedpoint of a Beverage is from ***dissipative losses*** in soil below the antenna, not from losses associated with radiation! In addition, termination resistance adds another source of loss (perhaps 30% of overall system loss). Overall, loss resistance in a Beverage is very high (several hundred ohms). This means the Beverage's feedpoint impedance is stable, even if mutual coupling radically changes radiation related impedances.

Note: Mutual coupling still remains, since mutual coupling is a function of element spacing and position. Even though mutual coupling still exists, it only affects the ***radiation resistance*** portion of element impedance. Since the large loss resistance (mostly from ground losses below the antenna) dilutes or swamps out mutual coupling effects, mutual impedances can be ignored.

Not only is a Beverage antenna's feedpoint impedance "immune" to mutual coupling effects when phased or placed near another Beverage, terminated Beverages offer a relatively constant feedpoint impedance over very wide frequency excursions. This makes arrays of Beverages ideal candidates for wide-bandwidth phasing systems, eliminating complex phasing and/or switching systems.

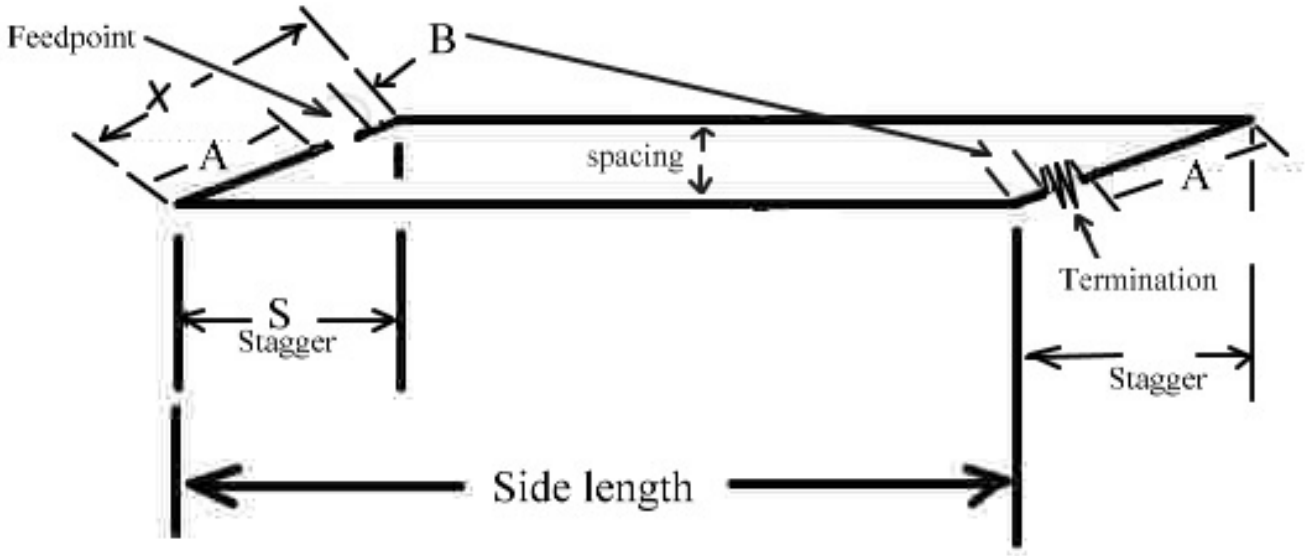
Types Of Phasing Systems

In order to better understand my antenna phasing systems, it is necessary to understand antenna phasing systems.

W8JI Parallelogram Array

Although it looks like a large horizontal loop, this antenna actually is two ground-connection-independent end-fire staggered (or stepped) Beverages. The short end wires form two single-wire feedlines at each end. One end-wire is series terminated with exactly twice the resistance of a normal Beverage, while the short wire at the opposite end is the feedpoint.

Stagger and spacing determines feed and termination location, the offset at the two ends being mirror images. A top view, looking down from straight above the antenna, looks like this:



This is an exactly "top down" view, looking down at the antenna from right above the antenna!

Click here for an Eznec file.

Notice the feedpoint is offset towards the termination end (front) of the antenna, and away from the null direction. This is typical for crossfire phased arrays. Crossfire arrays respond ***away*** from the delay line direction, exactly opposite conventional arrays. The termination is offset the same amount, but moves ***towards the feedpoint end*** of the antenna.

The feedpoint terminals, being floated (push-pull), provide 180-degree phasing between the two elements. The extra line length to the forward (left and front) element provides the "Stagger" delay.

Consider the actual wire length of a "short side" called "X" (which is the same as A+B). This length is the same on both ends of the antenna. The difference between A and B must equal or be slightly less than stagger (S).

To determine the offset of feedpoint and load:

1. Measure length of the end-wire, length "X".
2. Measure stagger in the end-fire direction, "S"
3. (X-S)/2=B
4. A=X-B

You may want to slightly offset the feed by making B longer and A shorter. This will move the null upwards, forming a cone. It is best to model the results.

Let's review a system, assuming S is 60ft and X is 100ft. We have X-S=40 divided by 2 for 20.

B is 20, A is then 80. The difference is 60, and that is the phase delay. Assuming 1.5 ft per degree that is 60/1.5 = 40 degrees delay. S is 60/1.5, or 40 degrees also.

We have a 180 shift at the push-pull feedpoint, so -40 rotates to +140. We have a 140 lead in the forward element, with a 40 degree spatial array delay. In the forward direction (towards termination) phase is (-40) + 140= 100 degrees out-of-phase. This results in nearly the voltage of one element alone, when the two element outputs of the long sides are summed. Towards the null (feedpoint) phase is +40 + 140 = 180 for a sin180=0 or zero voltage, a perfect 180 null.

On the second harmonic forward array feed system phase is -80 rotated to +100. Spatial array phase is now 80 degrees, or -80 towards termination. The result is (-80)+ 100= +20 in the forward direction. The result is nearly twice the voltage of one element in the forward direction. In the reverse direction, array phase is +80 + 100 or 180 degrees out-of-phase. We once again have zero back-fire response!

The general pattern holds true for any length of S less than 180 degrees, although grating lobes would make the pattern useless. The array, with 60ft stagger S, is usable from about 5 MHz down to VLF.

Summary, W8JI Parallelogram Array

Placing the feedpoint and termination centered on the end-wires is like using 180-degree phasing.

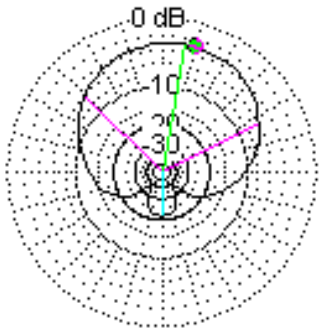
Placing the feedpoint and termination at the stagger minus side length distance is like using S-180 degrees phasing on any frequency with one exception, the antenna fires towards the feedpoint offset. This always results in a perfect backfire null, regardless of frequency.

Here is a wire table for a short 160-meter W8JI Parallelogram Array:

Wires

No.	End 1				End 2			
	X (ft)	Y (ft)	Z (ft)	Conn	X (ft)	Y (ft)	Z (ft)	Conn
1	0	0	6	W2E1	0	300	6	W4E1
2	0	0	6	W1E1	17	70	6	W3E1
3	17	70	6	W2E2	17	370	6	W4E2
4	0	300	6	W1E2	17	370	6	W3E2
*								

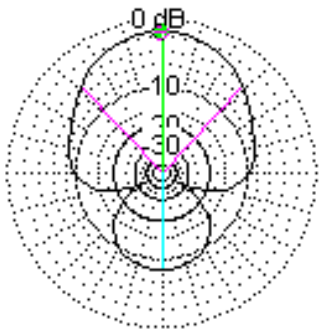
In a 370-foot length, this antenna has the following 20-degree wave angle pattern:



Azimuth Plot		Cursor Az	75.0 deg.
Elevation Angle	20.0 deg.	Gain	-18.92 dBi
Outer Ring	-15.99dBi		-2.93 dBmax

3D Max Gain	-15.99 dBi
Slice Max Gain	-18.91 dBi @ Az Angle = 80.0 deg.
Front/Back	19.2
Beamwidth	108.2 deg.; -3dB @ 26.9, 135.1 deg.
Sidelobe Gain	-37.54 dBi @ Az Angle = 270.0 deg.
Front/Sidelobe	18.63 dB

Average gain is -25dB, making the RDF about 6dB. In comparison a typical Beverage 370 feet long has this pattern:



Azimuth Plot		Cursor Az	90.0 deg.
Elevation Angle	20.0 deg.	Gain	-16.7 dBi
Outer Ring	-15.25dBi		-1.45 dBmax

3D Max Gain	-15.25 dBi
Slice Max Gain	-16.7 dBi @ Az Angle = 90.0 deg.
Front/Back	6.49
Beamwidth	84.8 deg.; -3dB @ 47.6, 132.4 deg.
Sidelobe Gain	-23.19 dBi @ Az Angle = 270.0 deg.
Front/Sidelobe	6.49 dB

If you wish to remove signals from the rear, the W8JI Parallelogram Array is a good broadband solution. It not only requires one transformer, one termination, and no ground systems at either end....it also is useful on several bands without switching anything!

Remember the above example is a reasonably short antenna, performance improves as side-length increases. There is no reason why this antenna can't be used from a band where it is only 1/2-wl long to bands where it is

several wavelengths long, as long as you properly choose stagger and width.

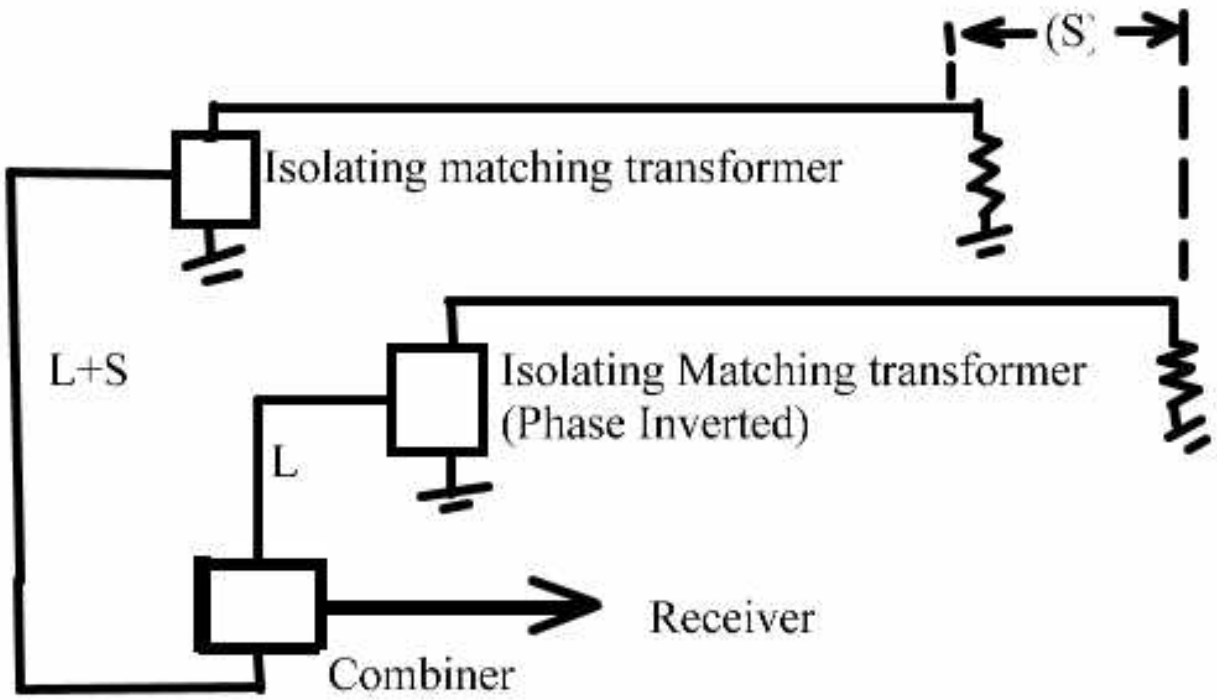
Crossfire-Echelon Beverage

This antenna is another frequency-independent antenna. It requires no band switching to work multiple bands. Because it uses cross fire phasing, rather than conventional phasing or hybrids, one phasing and delay line system covers several octaves of bandwidth.

A second feed method involves installing a pair of typical Beverages using isolation-type matching transformers, and inverting the phase of one antenna. Instead of being fed through two equal lengths of transmission line with a conventional splitter or combiner, the rearward stepped antenna (null direction) has an additional electrical delay equal-to or slightly-less-than the system's stagger distance in degrees. The delay line calculation can be done at any frequency, since stagger and delay change in step with frequency!

Performance is nearly identical to the above array, even very short antenna lengths (down to 1/2 wavelength) provide good F/B ratio. Of course the antenna will work better if length is increased. This arrangement is slightly more forgiving of dimensional errors than other systems, because each antenna is independent.

Remember you need a 180-degree flip at one feedpoint!! When you do this, the electrical delay of the antenna located towards the NULL must be S or slightly LESS than S. Using slightly less than S elevates the null and forms a cone, improving null usefulness and RDF of the array.



It is important to makes sure antenna feedpoint impedance is correctly matched to the transmission line. Be sure to measure SWR with an antenna analyzer!

You can also use a balanced wire feed, here is an EZnec model of an array using a combination W8JI parallelogram and Echelon-Crossfire feed.

Slinky and Loaded Beverages

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Slinky and Loaded Beverages

When space is limited and a Beverage antenna is too short for normal operation, increasing inductance (or capacitance) along the length of a "Beverage" will increase directivity. Inductance must be adding in series with the antenna, and doing so will increase the antenna's surge impedance.

Capacitance must be added in shunt with the antenna, and the extra capacitance reduces surge impedance. Additional shunt capacitance is very difficult to implement because each capacitor would need a ground, or you would need to surround the antenna with a slow velocity factor ***very*** thick dielectric. (Most of the air between the antenna and ground would have to be replaced with the dielectric.) There aren't any practical or easily implemented methods of slowing Vp to anything near the amount required through increased capacitance (at least none I could think of), so I will set that method aside.

There are three practical ways to slow velocity of propagation through inductance:

1. Use a spiral or zig-zag element
2. Use lumped inductor loading
3. A string of beads

How They Work

Contrary to what we might intuitively think, loaded antennas ***DO NOT*** work because we pack 800-ft of wire in a 200-ft area! They exhibit improved directivity because wave velocity is slowed, altering phase shift along the length of the antenna. Slinky (helical) and loaded Beverages show increased directivity because velocity of propagation along the antenna is ***decreased***.

As Vp is slowed, the antenna shows increased end-fire response and a gradually narrowing pattern. End-fire arrays fire in the direction of lagging phase, and the slower Vp causes more optimum phasing but only within certain limits. If phase is retarded too much, it actually starts to bring the signal back out-of-phase. With too much delay, the array tries to fire in the reverse direction.

Because of the reverse firing effect, there is a definite limit in phase delay (or Vp slowing) a system will tolerate. If the design goes beyond the optimum value, the antenna pattern tends to reverse direction, causing directivity to decrease. I've found optimum Vp for a 1/2 wl wire is generally around 0.5 times freespace velocity.

Slinky users should be particularly cautious to extend the coils an optimum amount. With 1/2 wl of distance, you would want somewhat less than 1wl of total conductor length. Too many "turns-per-foot" and the slinky, like any loaded Beverage, will try to fire backwards (towards the feedpoint).

Optimum termination impedance will always be somewhat higher than a conventional Beverage antenna. Surge impedance of the antenna is increased by the additional series inductance.

Helices vs. Lumped

For all practical purposes helices and lumped antennas are electrically identical. It doesn't matter at all if the antenna uses a string of soft-iron beads, a series of lumped inductors, zig-zags, stubs, or helices (a Slinky). The only criteria is the spacing between lumped loads must be a small fraction of a wavelength. For all practical purposes, 1/8th wl or less load spacing distance will make a lumped system perform identical to continuously loaded systems (slinkies).

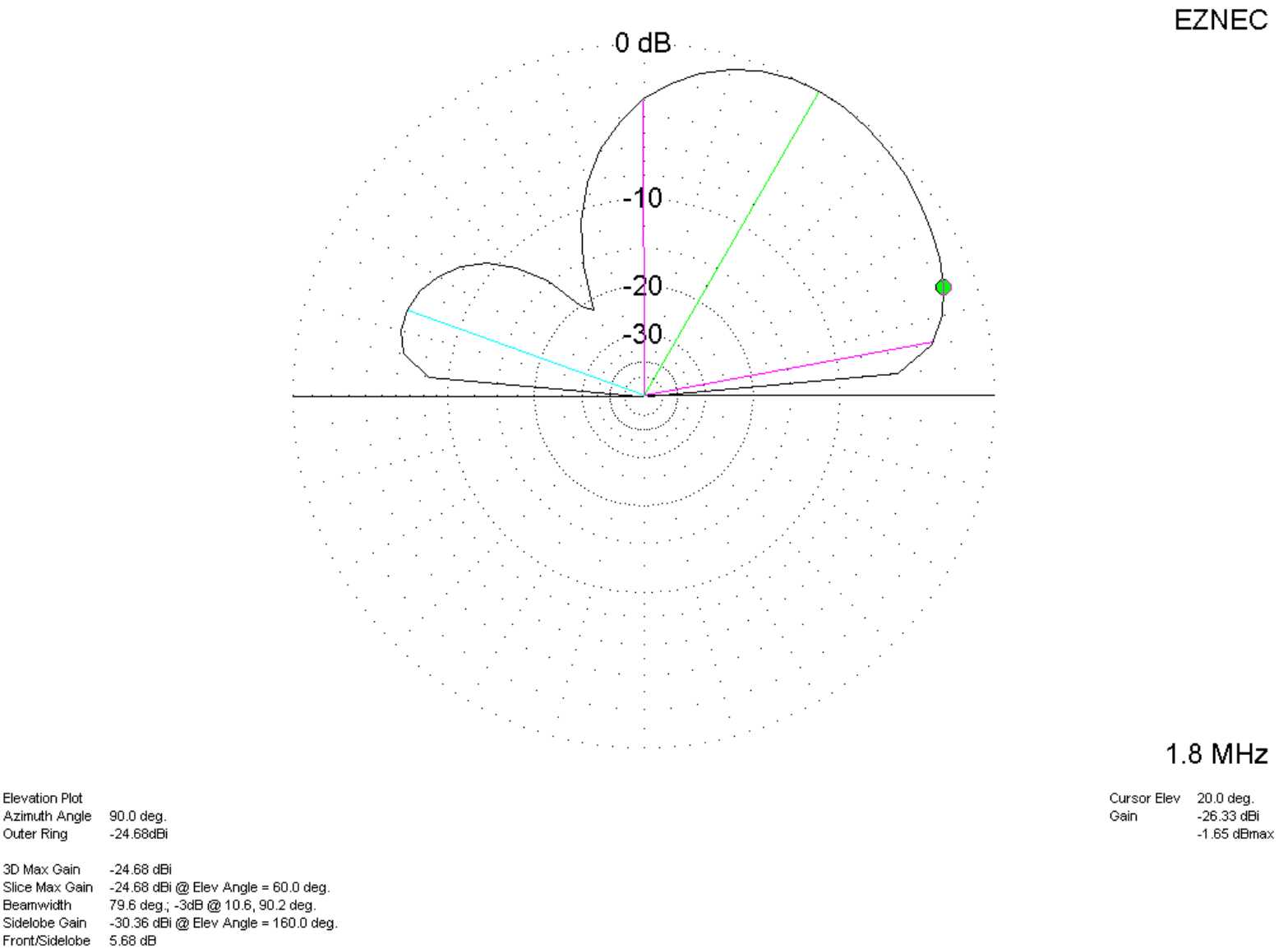
Lumped reactances, whether in the form of a bead, stub, or conventional coil should have at the minimum a somewhat modest Q. This is just another way of saying the inductances should not add excessive series resistance. ***Total*** reactance of ***ALL*** loads should be between two and fives times the surge impedance of the antenna. With nine inductors along a 1/2 wl wire, I've found optimum reactance to range between 150 and 250 ohms. Surge impedance was about 700-1100 ohms in systems I tested. A Q of 20 or more would be acceptable in each inductor.

Attached Model

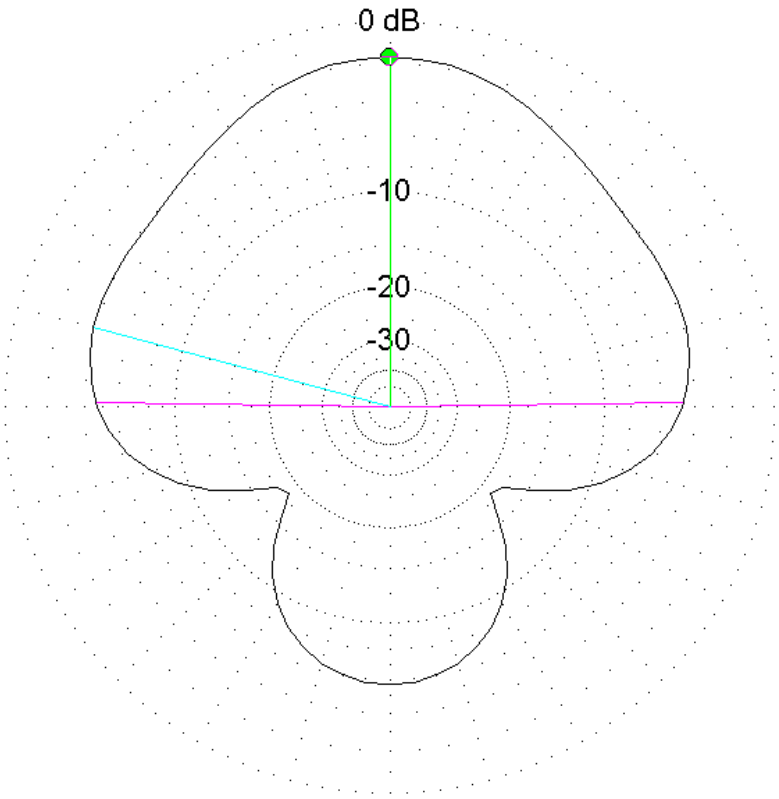
I've added an Eznec file of a loaded Beverage. A few things to note are I used radials to avoid a connection to high-accuracy ground at the termination. The feedpoint isn't critical, since it does not greatly affect directivity in this antenna. Unlike many antenna models, it appears a Mininec ground can be used without harming results.

Using a variety of ground-types I've found RDF remains at about 6dB. This places this antenna slightly below most elongated terminated loops, but well ahead of conventional short beverages.

loaded_bev



EZNEC



1.8 MHz

Azimuth Plot
Elevation Angle 20.0 deg.
Outer Ring -24.68dBi

Cursor Az 90.0 deg.
Gain -26.33 dBi
-1.65 dBmax

3D Max Gain -24.68 dBi
Slice Max Gain -26.33 dBi @ Az Angle = 90.0 deg.
Front/Back 4.03
Beamwidth 178.0 deg.; -3dB @ 1.0, 179.0 deg.
Sidelobe Gain -28.58 dBi @ Az Angle = 165.0 deg.
Front/Sidelobe 2.25 dB

Phasing Systems

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Standard Delay Line Systems

Many phasing systems use standard ***180-Spacing=Delay line*** phasing. In other words, the end-fire distance in degrees is subtracted from 180-degrees, and the result is the delay line length. This is an acceptable method in ***single frequency or narrow-band*** systems.

Conventional delay line systems have the following problems when used in endfire unidirectional arrays:

- Phase shift almost always generally changes the wrong direction as frequency is varied
- Phase shift only equals line length when the line is perfectly terminated or when the line is lossless and an exact multiple of 1/4wl
- Voltages or currents at both ends of the line are equal only when the line is perfectly terminated or an exact multiple of 1/2 wl and has negligible loss
- Currents at the load end of an odd-quarter wave (or multiple) line are only equal when the lines are fed from the same voltage and have negligible loss

Hybrids

Some systems use ***90-degree Hybrids or L/C phasing systems.*** Hybrids offer ideal distribution of power and provide the expected phase-shift only when perfectly constructed and terminated. While there is some tendency to self-compensate phase, they still suffer bandwidth limitations. Hybrids are very useful in electronic systems, amplifiers are one example. In an amplifier system, we might want a constant 90-degree phase shift despite slight frequency shifts.

Unidirectional antenna systems are never optimized when phase shift is fixed at some arbitrary value that remains constant as frequency is varied. ALL *unidirectional endfire* arrays require phase to track with element spacing change in degrees, as frequency is varied. Lumped component systems and Hybrids might save space, but they do not enhance array performance over the proper choice and design of transmission line phasing systems. This is true in both transmitting and receiving antenna systems!

If hybrids are so poor, why do we see so many of them in antennas? There are a few reasons

authors and manufacturers use hybrids. They often think:

- Hybrids assist in maintaining proper currents in elements
- The dump resistor only absorbs a portion of reflected power that would be wasted anyway
- Power is wasted only when the load SWR is high
- The association with Collins engineering must mean the system works well in any application
- Well-tuned Hi-Q components must work better than transmission lines

In reality, none of the above are true in a broad sense. Hybrids have their place, but it certainly is not in wide-bandwidth phased arrays.

Transmission-line phasing systems are a bit more tolerant than Hybrids. For example, a 90-degree long transmission line has zero degree phase error, even when grossly misterminated. A 90-degree phase delay transmission line has less phase error than a quadrature (90-degree) lumped component Hybrid when mismatched. Phase error peaks in misterminated transmission lines when the line is any odd-multiple of 1/8 wavelength, and is minimum with lines any multiple (even or odd) of 1/4-wl. For example, a 45-degree long transmission line provides 27 degrees phase lag when terminated in 25 ohms...not the 45 degrees people often expect! (Remember this when you see phasing designs that just throw a certain length of cable in series with a mismatched impedance!)

With all systems, amplitude errors are a problem. There isn't any passive system that provides correct phase ***and*** amplitude when load impedance changes, especially over a wide frequency range.

Crossfire Phasing

I prefer ***cross-fire phasing***, rather than the conventional narrow-band phasing methods discussed above. Cross-fire phasing, when designed properly, ensures phasing is always correct regardless of frequency. When elements (in this article Beverages) offer a near-constant impedance that is almost entirely resistive over a wide frequency range, cross-fire phasing can function perfectly from VLF to LF all in one antenna. Phase and amplitude can be designed for a back-fire null, with the upper limit in frequency set by element spacing and the lower limit set by array sensitivity. It is possible to design cross-fire receiving arrays that maintain the same basic directional response over several octaves of bandwidth.

Crossfire Phasing

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Crossfire Phasing

Most amateur antennas are copies of the ideas used in single frequency systems, like AM BC stations. Most of my work has been in HF systems in and around amateur bands. Generally these systems require wide bandwidth, rather than a certain pattern on one single frequency.

The proper way to extend bandwidth is to use cross-fire phasing, where one element's phase is rotated 180-degrees and the delay is set approximately at the spacing. Here's how it works:

We have two elements with spacing "s" at X and Y. The signals come from direction 1 or 2.

1-----X-----s------Y------2

Let's assume X and Y are 90-degrees apart on 80 meters.

A traditional array would delay the array feed to X and Y by 180-s = 90 degrees. Assuming X lags Y by 90 degrees, the phase of a signal arriving at X from direction 1 would be 0 degrees for the reference. The signal continues on to arrive at Y 90 degrees later because of the physical distance in space. Since X lags Y in the array phasing, the phase at the common point of the feed system is now X's phase of zero plus -90 delay is -90. The additional signal delay in space to Y is also -90. The signals are in phase, and the result is the in-phase sum of the two signals. Our receiving level has doubled over one element.

For a signal coming from 2, the phase of element Y is zero. Element X receives the signal 90-degrees later for -90 phase. That -90 adds to a -90 in the delay lines, and the result is -180 at the common point. The signals subtract exactly to zero, and there is no response.

Now assume we move the array to 160 meters, where our fixed length delay line and spacing are both 45-degrees long. The in-phase direction of 1 is still the direct sum of the two signals, but the out-of-phase direction becomes (-45) + (-45) degrees or -90 degrees at the common point. The signal vector addition is now unity. We have a very poor null!

Now let's go back to 80 meters and invert one element 180-degrees someplace in the system. In this case, we have:

From 1 we have X=0 degrees plus -90 though air is -90 at Y. Assume Y is inverted (it also works if we invert only X). The phase at Y is -90 but rotated to the opposite side of a "phase circle" by the mirror flip of 180 in element Y's inverted driving system. A -90 with a 180 flip is now +90. This combines with the -90 delay from element X through it's phasing line delay, and the sum of two equal signals of -90 and +90 is zero. -90 is exactly 180 degrees from +90, and the array now has no response towards 1.

When the signal arrives from 2, the phase is now inverted the same way and the result is perfect addition. We have Y =0 plus invert 180 is -180. Space delay is -90 plus -90 in the delay line for a net delay of -180. The result is signals from 2 are now in phase from both elements. The only change by adding a 180-degree element flip is the array inverts direction!

Here is where it gets interesting, and I am amazed so many amateur antenna designers miss this. Y is still inverted 180. When we move to 160 meters, the delay from direction 1 is:

X=0 The delay line adds -45, the same as the distance through space to Y. Since the delay line and space are the same, the delays effectively cancel. Since element Y is inverted, the combined phase at the common point is now exactly 180-degrees out of phase. the result is zero response towards 1!!!

More amazing, this is true for any frequency! There is always a perfect null towards 1 as long as the signal levels are even and impedances are matched from the two antennas, and we use a physical delay line rather than lumped components.

From direction 2 we have:

Y=0 plus a 180 flip for -180. At X we have -45 space plus the delay line of -45 for -90. We have -90 difference. The signal is the same as the signal from one element.

The only effect, as frequency is reduced, is sensitivity of the array drops. We have less signal, but we would have that anyway even if we used the narrow band phasing systems commonly used!

Why do amateurs, who almost always move around in frequency, use single-frequency 180-s phasing systems? Probably for the same reason we use 90 and 180-degree shift in a four square. We started out wrong, and just kept doing the same thing. While this won't provide octaves of bandwidth on transmitting, it does reduce phase errors across a single band substantially. It also provides a phasing method that allows us to build receiving arrays covering octaves without any loss of null direction or depth!

Combiner and Splitters

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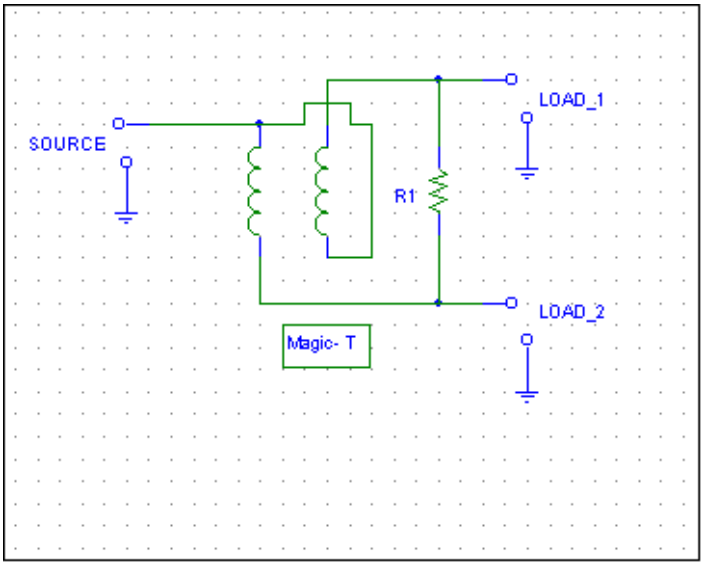
Combiners and Splitters

The magic "T" is a very useful device. It can provide equal voltages, equal current, or equal power to matched or unmatched loads. It is not a magic bullet.

The basic element looks like this:



For 0.1 to 30MHz applications, the transformer is a twisted pair of small (number 18 to 26 gauge) enameled wires with five to seven passes through a 73 material one-inch binocular core. The source impedance is parallel combination of the two load impedances, and generally is stepped back up through a 5:7 turn ratio (1:2 impedance ratio) transformer. R1 is twice the load impedance in splitter/combiner applications.

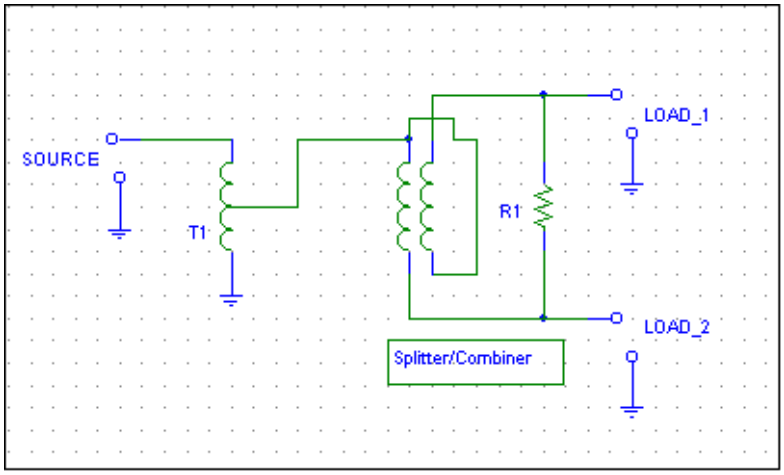
Like all passive splitters and combiners, this device is less than ideal. Isolation is only maximum when at least one port, and often two, are properly terminated.

This system does have an interesting characteristic, it can be used to force equal current into two different load impedances (or equal power by adjusting R1 to some value between zero and infinity). If R1 is open, each load is forced to have equal currents. If R1 is shorted, the loads naturally see equal voltages (T1 can be removed and the entire circuit configured as a T connector).

By forcing equal currents, a 1/2 wl transmission line can be used to feed a directional array instead of the more typical odd-1/4 wl line. With a 2:1 load resistance unbalance, I measured less than 0.1dB current error in such an application.

Receiving Splitters

Virtually all splitters are based on the Magic T with a matching transformer on the input. This splitters are certainly less than ideal, but they are simple and many times (but not always) better than a direct parallel connection.



T1 is a 7-turn transformer tapped at 5 turns (1.4:1 turns ratio, 2:1 impedance ratio) step-down transformer. 73 material binocular cores are ideal for 100kHZ to 30MHz.

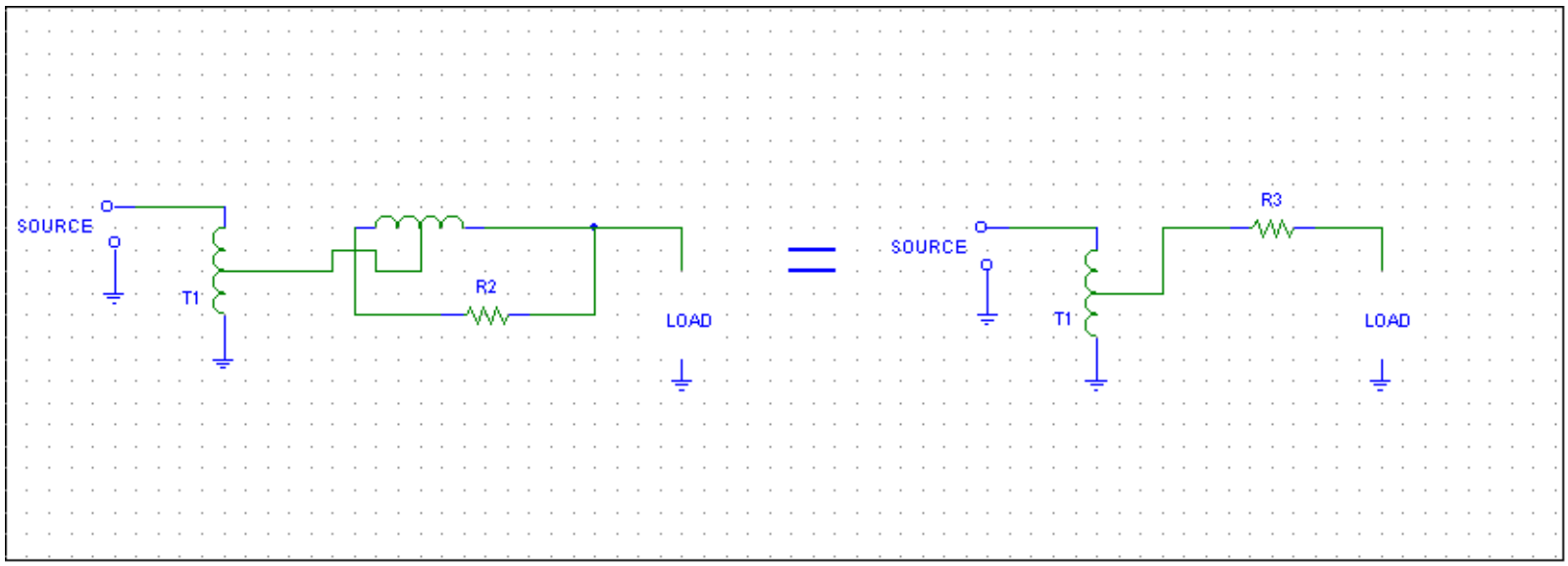
The magic T transformer is 5 to 10 turns of twisted pair wire through a 73 material binocular core. Configured as a center-tapped winding.

R1 is twice the expected load impedance. For 50- ohm systems use a 100 ohm resistor.

Losses

Typical loss of the Magic T is 3dB per port. The only other significant loss, when properly terminated, is a slight loss in T1. Normal loss of the entire system to either output port is 3 to 3.5dB. When combining in-phase signals, loss is only the loss in T1, there is no "combining loss".

When the Magic T splitter or combiner is open-circuited on one port, the electrical equivalent is:



R2, normally twice the value of any port, is stepped down by the impedance ratio of 4:1 in the Magic T transformer when a load port is opened. In a 50-ohm system, this effectively results in 25 ohms (R3) in series with the load impedance of 50 ohms. The resulting impedance of 75 ohms is doubled by T1 to 150 ohms. The equivalent input SWR, when open on one port, will be 3:1.

Loss will be 1.25dB from mistermination of the source by the 3:1 VSWR (assuming it is a 50 ohm source) with an additional loss of 1.7dB as power divides between R3 and the output port. The **theoretical open circuit loss (including transformer losses) is about 3 dB, the same as if a load were connected.**

The Magic-T behaves in a similar manner if one output port is shorted, with the exception R2 appears directly across the load. This is the equivalent of a 100-ohm resistor shunting a 50-ohm load. This impedance (33.34 ohms) is quartered by the center tapped winding of the Magic T

to 8.3 ohms. T1 steps this impedance up to 16.6 ohms. The result is, once more, a 3:1 VSWR.

This means signal level at port 1 theoretically does not change when port 2 is open or shorted, or anything in between.

Like many systems, these losses are based on source and load impedances being ideal. This is almost never the case, and the optimum value of R2 is often significantly different from being twice the load resistance. Because of that, our actual systems may show significantly more level change than the theoretical value (which is no change) when one port is incorrectly terminated.

If the system does not divide signal properly, the systems requires additional gain. This gain comes at the expense of lower IM performance. This is particularly true if the low-headroom amplifiers typical of amateur radio "preamplifiers" are used.

Even with high dynamic push-pull CATV transistor amplifiers, I've often found it necessary to optimize the splitting system. For example, stock my R4C's typically have ~20-ohm input impedances on 160-meters. Using R4C's on one port and higher-impedance receivers (or worse yet an open-circuit) on the other port results in starving the already low-sensitivity R4C receivers for signal (despite the use of a "splitter") if they are used in conjunction with higher sensitivity higher impedance receivers like FT1000's. The solution is to make every load (all the receivers) look like 50-ohms, readjust R1 to a new value, or use emitter-follower "active splitters".

I opted to make all of my receivers "look like" 50 ohms, but your choice may be different.

Amplifiers

Caution must be used in selecting amplifiers. Virtually all commercial amplifiers I have tested have very limited dynamic range. They are fine for amplifying high-loss antennas like Flags and EWE's, but terrible when connected to large Beverage systems. Adding an additional 6dB or more gain to the output of a Beverage just to overcome splitting losses (plus problems associated with uneven power division) almost always pushes these amplifiers "over the top". If you are going to use single-ended amplifiers with small low-power devices, be prepared to use multiple amplifiers (one for each receiver) and place them after the splitter system.

It is far better to buy or build one good amplifier, but it should be a very high dynamic range device!

K9AY_Flag_Pennant_Ewe

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Elongated Terminated Loops

Terminated loops have only recently become popular, even though such systems were available commercially in the 1960's and earlier. Sometime in the early 1970's, I received an advertisement from the late W1BB. That advertising brochure described arrays of receiving loop antennas marketed by a Canadian manufacturer for commercial and military use. The brochure had pictures of diamond-shaped terminated loops, individual small loops, and loop arrays that looked like a number of Flag antennas strung end-to-end where each rectangular "loop cell" fed the next loop. The brochure gave directional patterns of the arrays, and the patterns looked much better than those of other antennas. Unfortunately the price, as I recall, was in the tens of thousand dollars for even the least expensive array! Although detailed technical information was withheld, it wasn't difficult to figure out the electrical design of each array.

That early brochure started me thinking in terms of antennas other than Beverages. I experimented with small loop antennas, and eventually installed an array of eight in-line loop cells. That system allowed me to be the first eastern USA station to work Japan and other Asian countries through interference from multiple multi-KW LORAN transmitters, and as the system was refined I could work JA's on a regular basis long before the broad LORAN pulse transmitters were removed from 1900 kHz. This is a testimony to the clean pattern and wide deep nulls possible using arrays other than traditional Beverage antennas.

Are they quiet because they are loops? Not at all!

One of the most damaging rumors to good antenna science are claims that loop antennas have a mystical property that somehow rejects noise. Nothing is further from the truth, with one exception. Corona.

Loops do not have sharp points sticking out into air at high impedance points, and that can be an advantage in limited cases.

Any sharp point extended from the area of anything, even wooden ship's masts in old sailing vessels, is more subject to corona in inclement weather. Sailors feared St. Elmo's Fire long before anyone knew what electricity was, and amateurs fear it today when trying to receive. We often just don't know what to call it, and can't see it in our brighter skies, so we call it "precipitation static".

The sharp point in a dipole where the wire or tubing ends not only promotes corona by virtue of the

fact it is "out in space by itself", it almost always has the disadvantage of being at a high impedance point in the system. The tiny random charge movements with little current and very high voltage (high impedance source) inherent in conductors around the corona discharge are driving a high impedance part of the system, an ideal situation to maximize transfer of tiny amounts of noise power into the system!

That is almost certainly where the rumor that loops reject noise comes from. We forget it is a function of the sharpness and impedance in the area where the antenna is sharply shaped, and instead fixate on the fact it is "a quiet loop".

Are terminated loops actually loops?

Terminated loops do not behave like small conventional loop antennas. They do not carry uniform in-phase currents throughout the area of the antenna. They don't behave like directional couplers either, they are antennas...not nearfield coupling devices coupled to other conductors (we hope). Terminated loops are really just short verticals, with the phasing inherent in the long horizontal wires. The loop's vertical areas receive the desired signals, while the horizontal conductors merely serve to act as transmission/phasing lines for the vertical (or sloping vertical) ends. A small elongated terminated loop acts like a simple two-element vertical array with integrated phasing lines.

This is true even in the K9AY Loop and Pennant antennas, which have sloped conductors. The sloped conductor behaves as a vertical (think about this when people follow the mistaken advice that sloping the last few feet of a Beverage "stops noise pickup of vertically polarized signals"), ten feet of vertical drop is still ten feet of vertical conductor exposed to vertically polarized signals. The fact the antennas work so well is testimony to how sensitive a sloped wire is to vertical polarization. If sloped wires weren't sensitive to vertical signals, the Pennant and K9AY Loop wouldn't work!

Elongated terminated loops, and arrays of elongated terminated loops, are a special form of short phased verticals where phasing and feed systems are an integral part of the antenna. This integrated system of elements, feed system, and phasing solves construction problems associated with arrays using more recognizable vertical elements. With any system there are tradeoffs, elongated loop antennas only allow very limited control of phase and unwanted high-angle response. We can't obtain optimum patterns (although a properly placed series capacitance will help) because this system only allows limited control of current distribution and phase shift.

These antennas "want to" be verticals because the earth below them and somewhat closer spacing prevents the horizontal components from being efficiently coupled to space while the earth simultaneously enhanced the vertical radiation component of the gradually sloped conductors making up the antenna. The horizontal sections act as a wide-spaced air dielectric transmission line.

These antennas work best when propagation time of signals along the horizontal wires matches propagation time of the wave in space around the antenna, and when the earth or a ground system below the antenna is good. Suppression of high-angle horizontally polarized radiation, and maintaining velocity of propagation near unity, are why EWE antennas (and other forms of this family of antennas) thrive on good ground systems or good earth below the antenna. The desired earth effects are opposite those desired with Beverage antennas, an ideal situation for highly conductive soil!

Why use terminated loops?

It is easy to understand why "loop" systems, even very small loop systems, have become popular. Arrays of terminated loops vertical elements produce effective low-angle receiving performance along with a relatively clean pattern. EWE's, Flags, Pennants, and K9AY loops are effective methods of building two element vertical arrays. They are easy, small, and inexpensive! They are noticeably less directive than two phased verticals would be, because the horizontal components are not totally cancelled by ground effects and the opposing wire, but they certainly are easier to construct than phased verticals (with all their loading, grounds, and coaxial cables). It's all the standard old tradeoff we just can't seem to get away from, we always must balance complexity against performance.

Many people are working with the various arrays of elongated loop antennas, so there are few contributions I can make other than describing how or why they work. I would like to suggest it is possible to extend the arrays end-to-end for some distance without external feed systems, and well-placed reactances can be used to modify patterns. Very little work has been done in that area. I'd suggest experimenting with series capacitors, perhaps placed mid-way in phasing (the horizontal wires) areas to increase velocity of propagation through elongated loop arrays and increase directivity.

Are small loops hyper-sensitive to vertical masts?

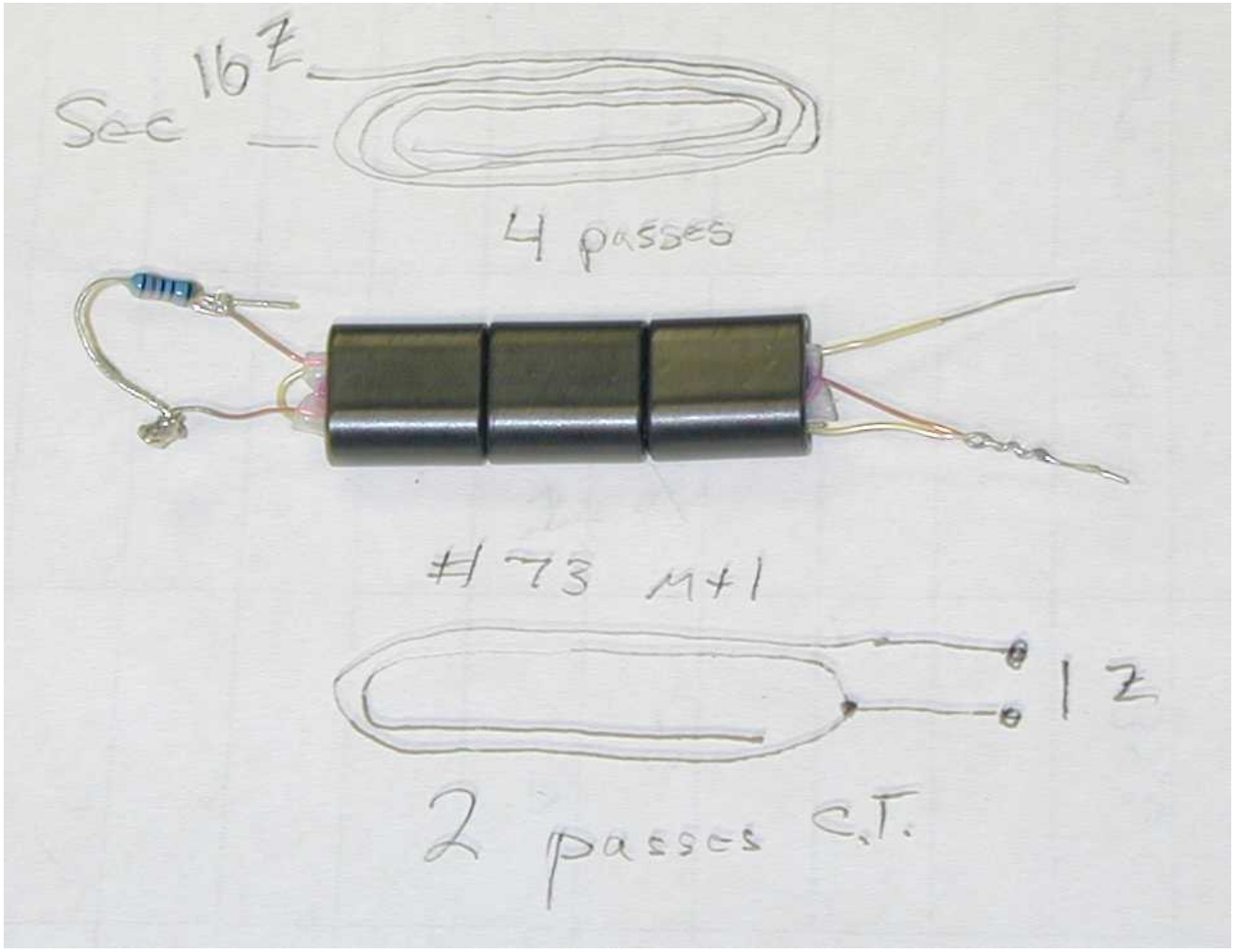
There is no compelling evidence that any of these antennas are more sensitive to vertical metal masts than any other antenna would be. As a matter of fact, the only basis for such claims appears to come from models that fail to pass simple recommended tests for model accuracy and stability. If we build a model that is flawed and oversensitive to changes in things like the number of segments used in the model, we can expect it to be sensitive to nearly any change!

Everyone is free to say and do what they like, but other than keeping a short mast a few feet away from vertical wires and NOT connecting that mast directly to the antenna or feedline, I wouldn't hesitate for a second using metal supports. My large arrays of loops in the early 70's used metal masts, my arrays in the 80's did, and as have commercial arrays.

Feedpoint Matching

Great care must be taken in decoupling the feedline from the antenna in the balanced versions of these antennas, although the EWE (being unbalanced) is relatively immune to such problems. Keep in mind the antenna generally looks capacitive as a common-mode structure, so inductive decoupling (i.e. a choke coil of coax) can actually ***increase*** system problems. The best common-mode isolation system would be an isolated winding transformer designed with minimal capacitance between the antenna winding and the rest of the system. I use a small transformer with stacked 73 material binocular cores in feeding some of my high impedance "log-Beverage" arrays, and similar transformers should work with ~1000 ohm impedance elongated loop antennas.

Because this transformer only has a single turn primary (two turns with the balancing pass), I'm able to reduce stray capacitance to a dozen pF or less. It has excellent balance, low SWR over a wide bandwidth, and very low loss. The reasonably low transformer capacitance, when used in concert with proper feedline grounding and routing, should make the system relatively immune to common mode problems. I'd route the feedline horizontally directly away from the end of the antenna for a few dozen feet (but never a distance approaching 1/10th wavelength or longer), and then drop the feedline to ground earthing the feedline shield at that point. Decoupling beads or sleeves belong on the receiver side of the shield grounding point, not between the ground and the antenna!



I wind this transformer on three Fair Rite Products 2873000202 cores (about 1/2 inch square and 1/3 inch thick 73 material). The high impedance (secondary) winding is #26 enameled wire through Teflon tubing, while the primary is Teflon coated wire-wrap wire wound outside the tubing. The small extra pass that "dangles" on the low-Z primary winding helps balance the system, even though it adds a few pF of capacitance.

By the way, a Faraday shield will only make things worse. It will increase unwanted stray capacitance and might deteriorate the high impedance winding's balance if it is not properly grounded. The proper grounding point for a Faraday shielded primary is *opposite* the exit point of the primary winding, or on the secondary winding's exit side of the transformer. Most Faraday shields described for Beverage and other transformers are not only useless, they are often incorrectly grounded and actually increase unwanted coupling!

Eznec file courtesy of Roy Lewallen.

K9AY_Lewallen

Receiving_basics

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Receiving Antenna Design

My local wintertime 350Hz BW noise (after amplifier) compared to a sample of signals on one night was:

Noise -127dBm

9H1BM -122dBm

OM0WR -95dBm

DF2PY -88dBm

WA8OLN -78dBm

W3GH -60dBm

W4ZV -32dBm

The dynamic range between noise and W4ZV was 95dB! It also illustrates how important antennas, location, and propagation are rather than power. The dB difference between signals from the same area is profound. Many signals run at or near noise floor. This may not be typical of every night, but it shows how large the signal level variations between weak DX and strong signals are.

Before talking about receiving antennas for lower frequencies, it is important to understand a few basics. We all understand the primary reason we use special receiving antenna systems is to improve signal-to-noise ratio. On the surface this sounds like the same reason we use directional transmitting antennas, but there are some very important differences between transmitting and receiving applications.

One thing we often hear is how a 7dBd gain yagi will improve receiving by a similar amount over a dipole, the logic behind that idea is the gain works both ways. I've read that time and again as justification for using a high gain transmitting antenna. On the surface, it does sound reasonable and correct. Yet it's really not correct in most (if not all) cases! A 7dBd gain antenna with 20 dB nulls can improve receiving an amount from nothing to 30 dB or more, depending on many things we often don't consider.

When transmitting, we can not tolerate much power loss. The most important antenna characteristic, other than a reasonable SWR and having the antenna remain in the air for a while, is having maximum possible field strength in a given direction or directions. In general, the single most important parameter is having maximum gain in all the desired directions and at the most useful wave angles. We know, for a given polarization and wave angle, the highest gain antenna will deliver the strongest signal to our target area. Transmitting antenna choice is reasonably simple, we want properly positioned gain. Take-off angle, side-lobe strength, and F/B ratio take a back seat to maximum gain at the desired wave angle and direction. We can generally model the antenna and confirm the results with a few simple

measurements, and be confident we have made a good choice. By the way, if you hear anyone selecting an antenna based on take-off angle help him out. Take off angle is not important, what we actually need is maximum possible gain at the desired angle and direction. After all, we don't care where the peak is as long as the antenna we pick has more signal (gain) at the desired spot than other antenna choices!

When receiving, the system has a much different design priority. If we consider gain, we can be totally misled. It's easy and somewhat common to make poor choices when picking receiving antennas, mostly because we often don't know what to look for. What we look for can also be different depending on the actual application. A system that works extremely well in some applications can be a real dud in other applications. I'll explain how to make the best choice and what the pitfalls are, but first we have to understand some basics. ***If you would like to hear a demonstration of how excellent directivity reduces noise, listen to QRN_JA2ZJW on the DX Signal Recording page.***

Antennas and Fields

If we are going to pick the best receiving antenna, it is important we understand how antennas respond to "signals" and "noise". Unfortunately, this is probably the single most misunderstood part of receiving antenna design. This a problem mostly because we are constantly bombarded with folklore giving us the wrong impression of how antennas work.

One example that comes to mind is small loops, often referred to as "magnetic radiators". Folklore often considers the radiation or response of these antennas as somewhat magical, with the antenna behaving like a sieve that sorts "good magnetic signals" from "bad electrical noise". Nothing is further from the truth! At a relatively small distance the small magnetic loop is more sensitive to electric fields than a small electric field probe. The ratio of electric to magnetic fields are sometimes called the "field impedance" of the antenna. An antenna with a high field impedance has a dominant electric field, and an antenna with a low field impedance has a dominant magnetic field, when both are compared to freespace ratios at a large distance from the antenna.

At distances of about $\lambda/2$, the field impedances of all by physically large antennas are almost indistinguishable from each other, no matter what the antenna type. Strange things happen at closer distances, as we will soon see. Once we understand the basics, we might picture our antennas differently. We might find antennas don't respond to signals and noise like we thought, but at least we won't be as susceptible to the "noise" caused by rumors and folklore!

Induction field

The induction-field area involves the energy storage area around the antenna, and involves both electric and magnetic fields. These energy storage fields dominate the system near the antenna, although the radiation still exists. Contrary to some bizarre claims (i.e. CFA or CTHA Antennas), there is no possible way to "mix" induction fields and create a radiation field. The effects that cause each are different, and they follow different rules. The induction fields fall off much more rapidly with distance than radiation does, even though we "measure" or talk about each in terms of magnetic and electric effects.

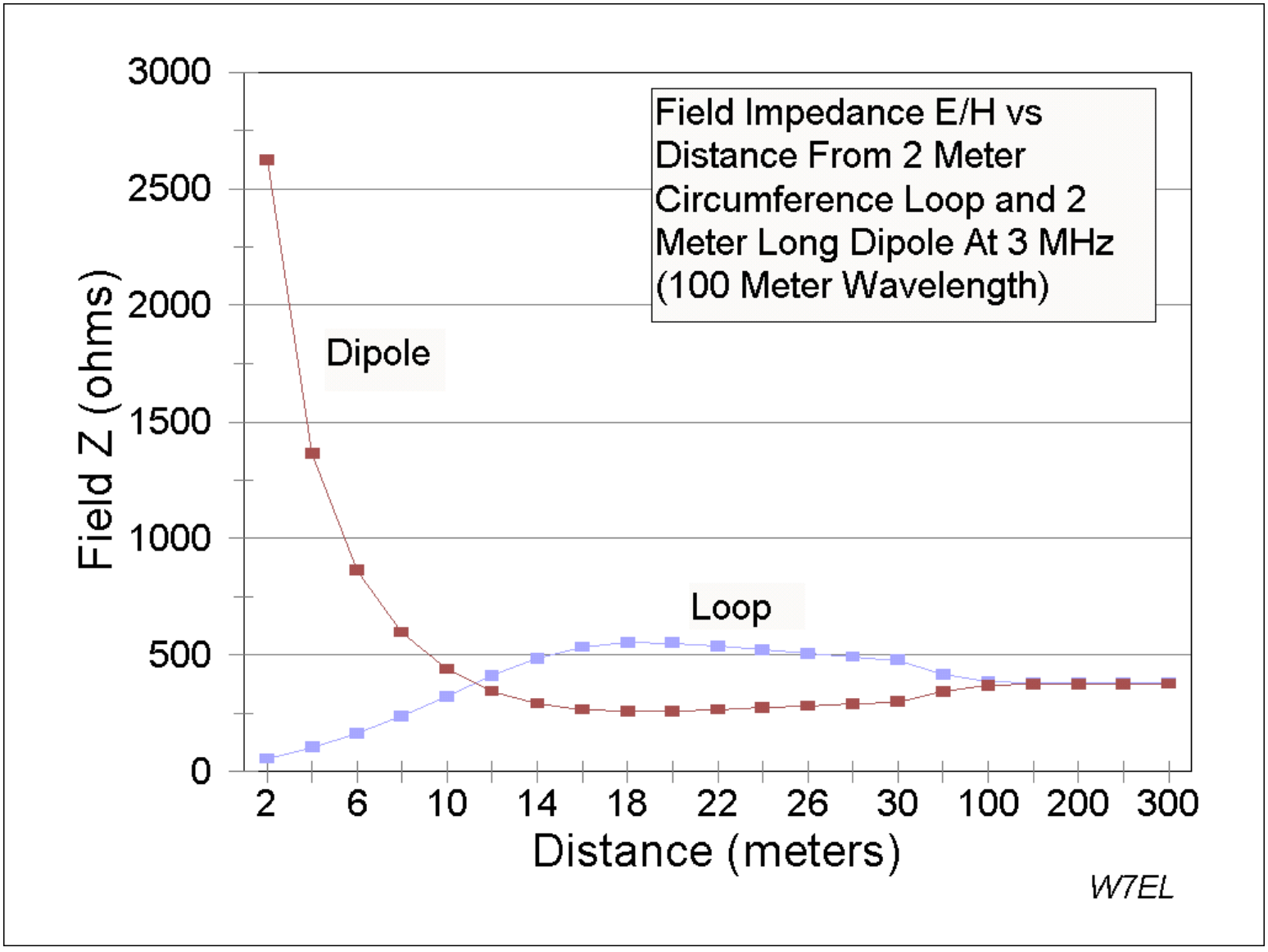
We can move the fields around, concentrate them or spread them out, and even change the ratio of these fields...but every antenna has both electric and magnetic induction fields. ***Whenever we have a time-varying electric field, it must be accompanied by a time-varying magnetic field.***

We might consider induction fields part of direct capacitive or inductive coupling between areas of the antenna and objects or space around the antenna. Close to the antenna, field intensity might not fall off as the rules normally predict with distance. Very near an antenna, fields are an accumulation of effects from sources at multiple directions and distances. It isn't always easy to understand what actually is happening very near an antenna, where the area of the

antenna is large compared to the distance where we are observing the effects. Near the antenna, pattern and field impedance is generally nothing like we might intuitively imagine.

It is the response in this area, generally within λ/10 distance from the antenna, that small "magnetic loop" and "electric dipole" antennas get their names.

Very close to a small loop antenna (but not necessarily very near the open ends), the magnetic field dominates. We could describe that effect by saying say the field impedance is "low". Conversely, near a small dipole or monopole the electric field dominates, the antenna has a high field impedance. Remember, this generally applies only with λ/10 distance from the antenna.



The since the distance of a wavelength in the above graph (thanks W7EL) is 100 meters, we can also considered the bottom scale as a percentage of a wavelength. We can see at about 11 percent of a wavelength (which would be about 50 feet on 160 meters), there is no difference in field impedance between a small loop and a small dipole. At distances beyond 50 on 160 meters, the loop actually has a ***higher*** field impedance than a dipole.

So much for the myth that a receiving antenna can sort good signals from bad signals (noise) by virtue of being "magnetic"! We not only don't have the response we might imagine, we probably have no idea if the close-by unwanted signal source is radiated from a source that is electric or magnetic field dominant. Successful noise reduction by virtue of

by antenna "style" would mostly be a matter of hitting a lucky combination through careful experimentation.

Near Field

The near field area is an area where the ultimate pattern is not fully formed. It is possible, with large arrays of small elements, to be out of the induction field region but still well within the near field area. Let's consider individual groups of elements as "cells", and the array a combination of small directional cells occupying a very large physical area.

This occurs in some of my broadside/end-fire combination arrays because physically small antennas forming individual end-fire cells are combined forming an array more than one wavelength across. The array is physically large, and at a distances of several wavelengths the combined field clearly originates from multiple sources that vary in distance or direction. The pattern is not perfectly formed because the signal from closer elements is less attenuated by distance, and the phase difference between the cells at a given azimuth is not the same as the ultimate phase further away at the same azimuth.

This applies to utility line radiation, where distant noise sources are conducted along the entire line. To the receiving array, it might appear that a single arc on an insulator a mile away is actually radiating from multiple directions with many phase relationships. The field the receiving antenna "sees" might appear to arrive from multiple directions, making it difficult or impossible to null the noise with a simple directional cell or more complicated array of directional cells. This would be a good application for a device like an antenna noise canceller, because the noise at the antenna terminals will be the vector sum of response to multiple radiated fields reaching the receiving antenna from a single noise source. In this situation a sample of noise taken near the power line could be combined out-of-phase with the sum of radiation points along the power line to null the noise.

Clearly a noise source or sources might also be outside the induction-field dominant area, and still not have a fully formed pattern. Antennas must be in the far field of the noise source, and the noise source must be in the far field of the antenna if patterns are to be predictable when nulling noise. Besides the fact larger distances provide greater attenuation of noise, it is also very desirable to locate our receiving antennas in the far field of noise sources, and noise sources in the far field of the antenna. This makes a good case for placing receiving antennas as far from other conductors (such as power lines) as possible.

This doesn't means urban or suburban dwellers should give up and move to rural areas, it just means the results will be less predictable and actual experimentation is likely required. Results aren't always predictable, unless the antenna is a few wavelengths from other potential unwanted signal radiators or objects that might affect antenna pattern.

Noise

The noise that limits our ability to hear a weak signal on the lower bands is almost always an accumulation of many signal sources. Below 18 MHz, the noise we hear on our receivers (even at the quietest sites) comes from terrestrial sources. Receiver noise is generally a mixture of local groundwave and ionosphere propagated noise sources, although some of us suffer with dominant noise sources located very close to our antennas.

Our locations fall into three basic "radio" categories that may or may not be related to our actual communities:

Note: noise levels quoted in this text are the average of three independent studies by Bell Labs, FCC Land Mobile Advisory committee, and the Institute for Telecommunication Sciences. Rural data are actual measurements of summer noontime and winter midnight noise at my location, several miles from high voltage transmission lines and far from any industrial or suburban populations.

Urban

In urban-type noise situations, noise arrives from multiple random sources through direct and groundwave propagation from local sources. One or more sources can actually be the induction-field zone of our antennas (in most cases the induction field dominates at distances less than 1/2λ). Urban locations are the least desirable locations because typical noise floors average 16dB higher than suburban locations. There is often no evidence of winter night noise increase on 160 meters, since ionosphere-propagated noises are swamped out by the combined noise power of multiple local noise sources. Much of the noise sources are utility distribution lines, because of the large amount of hardware required to serve multiple users. Other noise sources are switching power supplies, arcing signs, and other unintentional man-made noise transmitters.

Suburban

Suburban locations average about 16 dB quieter than urban locations, and are about 20 dB noisier than rural locations. Noise generally is directional, arriving mostly from areas of densest population or the most noise-offensive power lines. Utility high-voltage transmission lines are often problematic at distances greater than a mile, and occasionally distribution lines can be problems. The recent influx of computers and switching power supplies has added a new dimension to suburban noise.

There is often a small increase in nighttime winter noise at exceptionally quiet suburban locations. This increase occurs when propagated terrestrial noise equals or exceeds local noise sources.

Rural

Rural locations, especially those miles from any population center, offer the quietest environment for low-band receiving. Daytime 160 meter noise levels are typically around 35-50 dB quieter than urban, more than 20 dB quieter than suburban locations. Nighttime brings a dramatic increase in low-band noise, as noise propagates in via the ionosphere from multiple distant sources.

Primary local noise sources are electric fences, switching power supplies, and utility lines. I can measure a 3 to 5dB daytime noise increase in the direction of two population centers, Barnesville (population 7500, distance 6 miles) and Forsyth (population 10,000, distance 6 miles) Georgia.

Typical daytime noise levels, measured on a 200 foot omni-directional vertical, are around -130 dBm with a 350 Hz bandwidth (noise power is directly proportional to receiver bandwidth) . Noise power increases about 5 to15 dB at night, when the band "opens". As in the case of suburban systems, directional antennas reduce noise power.

Nighttime is the "big equalizer", reducing the advantage of location as distant noises increase with improved propagation.

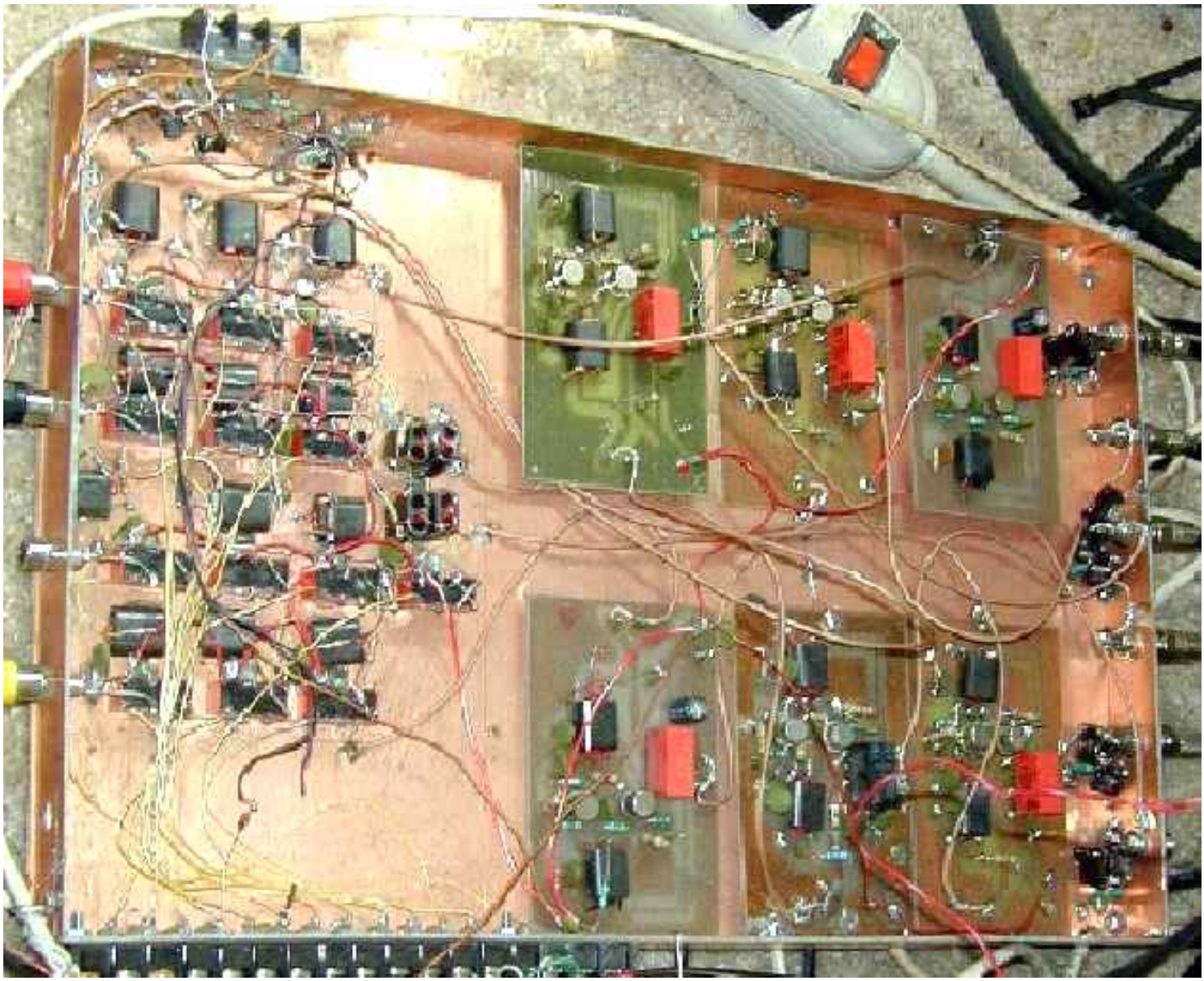
How Important is Gain?

When choosing one receiving antenna over another, never select an antenna by comparing antenna gain. Gain only affects receiving performance when the receiving system noise floor is determined by internal system noise. If we hear a definite decrease in noise when replacing the antenna with a dummy load of the same impedance, gain by itself is not a factor.

Noise not only varies with the hour of day, it also varies with direction and season. Further complicating matters, noise power is directly proportion to receiver bandwidth. Going from a 2.5kHz filter to a 250 Hz filter with the same basic shape-factor reduces noise by 10 dB. When determining if system gain (for a given directivity) is adequate to establish system noise floor, tests must be made at the quietest operating time using the narrowest selectivity. In general, we should set system gain to allow a comfortable increase in noise when the antenna is connected at a time when propagation is poor but adequate to allow some contacts. This sometimes results in excessive gain when propagation is more favorable, but it is simple to remove gain (either by adding an attenuator or removing an amplifier) when using wider bandwidths or if conditions are noisier.

My receiving antennas have high dynamic range amplifiers using push-pull 2N5109 transistors. Amplifier gain is approximately 15 dB, and noise figure is approximately 5 dB. Gain compression occurs at approximately 30 dBm (1 watt) output. The blocking dynamic range of the amplifiers (using a 250 Hz bandwidth on the receiver) is around 180 dB, making my receivers the limiting factor. The amplifiers have internal bypass relays, allowing them to be easily bypassed when required by removing amplifier power.

Each of these amplifiers drives a four-way splitter, allowing any one-of-four receivers to connect to any antenna. Multiple receivers can select the same antenna simultaneously. Splitter system loss is 7 dB. Typical net system gain is around 8 dB to each receiver. Some amplifiers are designed to have slightly more gain and some less gain, because some antenna systems have a lower noise floor than others.



What Antenna Parameter is Important?

Once any receiving system limits on *external* noise, antenna pattern becomes the only thing that matters. Here are the patterns of two Beverage antenna systems, each with different gains. One antenna uses two side-by-side Beverages, while the other is a single Beverage. I'll include Eznec wire tables so you can model the antennas, and see the mistake we often make.

First, the wire table can look like this:

Wires

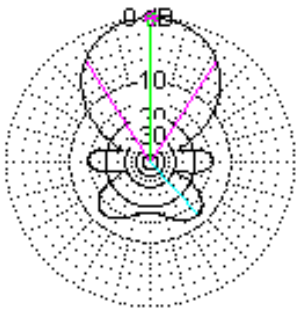
Wire Other

Coord Entry Mode Preserve Connections

Wires

No.	End 1				End 2				Diameter	Segs
	X (ft)	Y (ft)	Z (ft)	Conn	X (ft)	Y (ft)	Z (ft)	Conn	(in)	
1	0	0	0	Ground	0	0	7	W2E1	#16	5
2	0	0	7	W1E2	0	780	7	W4E2	#16	50
3	70	0	7	W5E2	70	780	7	W6E2	#16	50
4	0	780	0	Ground	0	780	7	W2E2	#16	5
5	70	0	0	Ground	70	0	7	W3E1	#16	5
6	70	780	0	Ground	70	780	7	W3E2	#16	5
*										

This is two Beverages 70 feet apart. If we terminate both of them and set the source of one at zero (so only one Beverage is active), the following pattern results:



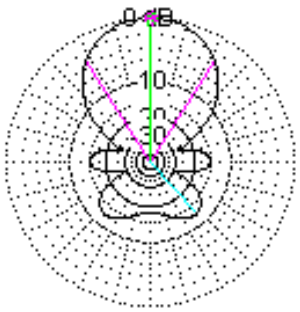
EZNEC

1.83 MHz

Azimuth Plot		Cursor Az	90.0 deg.
Elevation Angle	20.0 deg.	Gain	-11.24 dBi
Outer Ring	-11.24dBi		0.0 dBmax

Slice Max Gain	-11.24 dBi @ Az Angle = 90.0 deg.
Front/Back	18.11
Beamwidth	65.7 deg.; -3dB @ 56.8, 122.5 deg.
Sidelobe Gain	-23.33 dBi @ Az Angle = 312.0 deg.
Front/Sidelobe	12.09 dB

Now we add the second source, so both parallel Beverages are active. The pattern and gain is now:



EZNEC

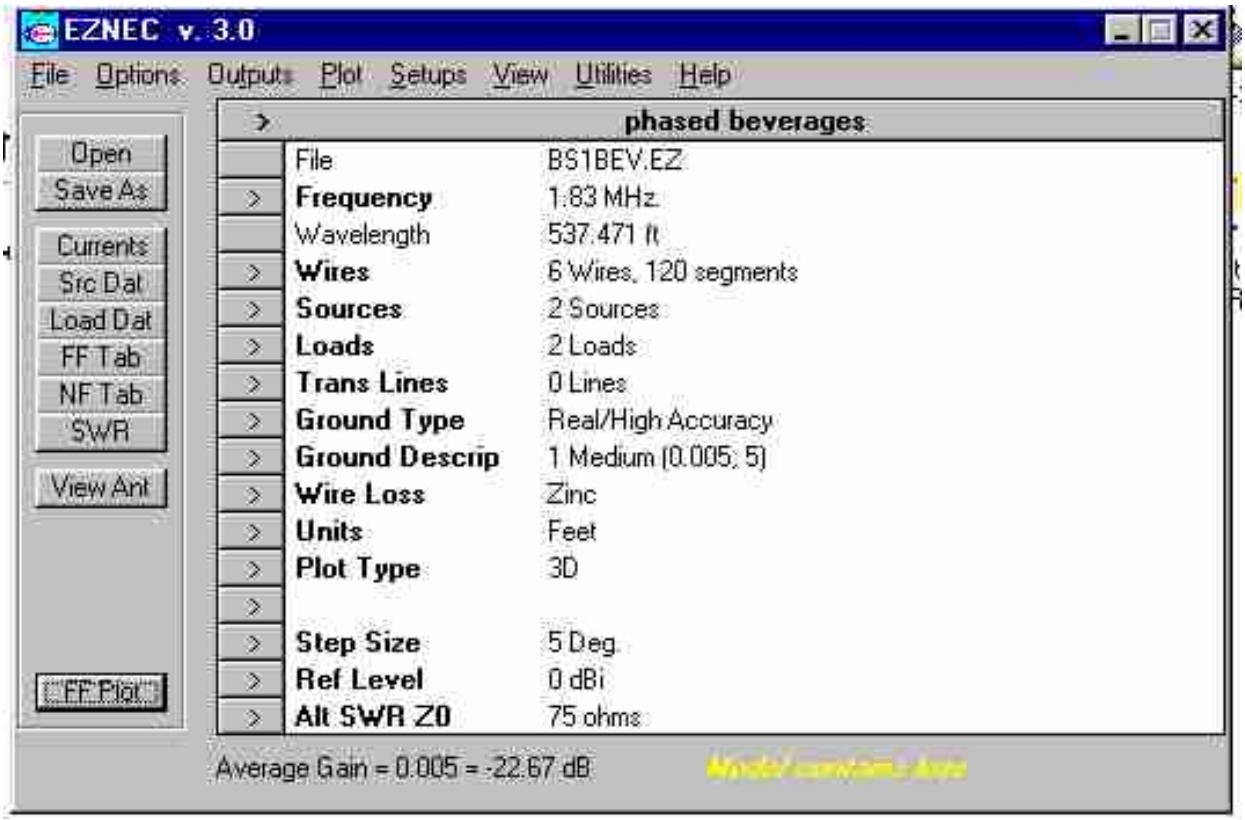
1.83 MHz

Azimuth Plot		Cursor Az	90.0 deg.
Elevation Angle	20.0 deg.	Gain	-8.2 dBi
Outer Ring	-8.2dBi		0.0 dBmax

Slice Max Gain	-8.2 dBi @ Az Angle = 90.0 deg.
Front/Back	18.11
Beamwidth	63.8 deg.; -3dB @ 58.1, 121.9 deg.
Sidelobe Gain	-20.76 dBi @ Az Angle = 312.0 deg.
Front/Sidelobe	12.56 dB

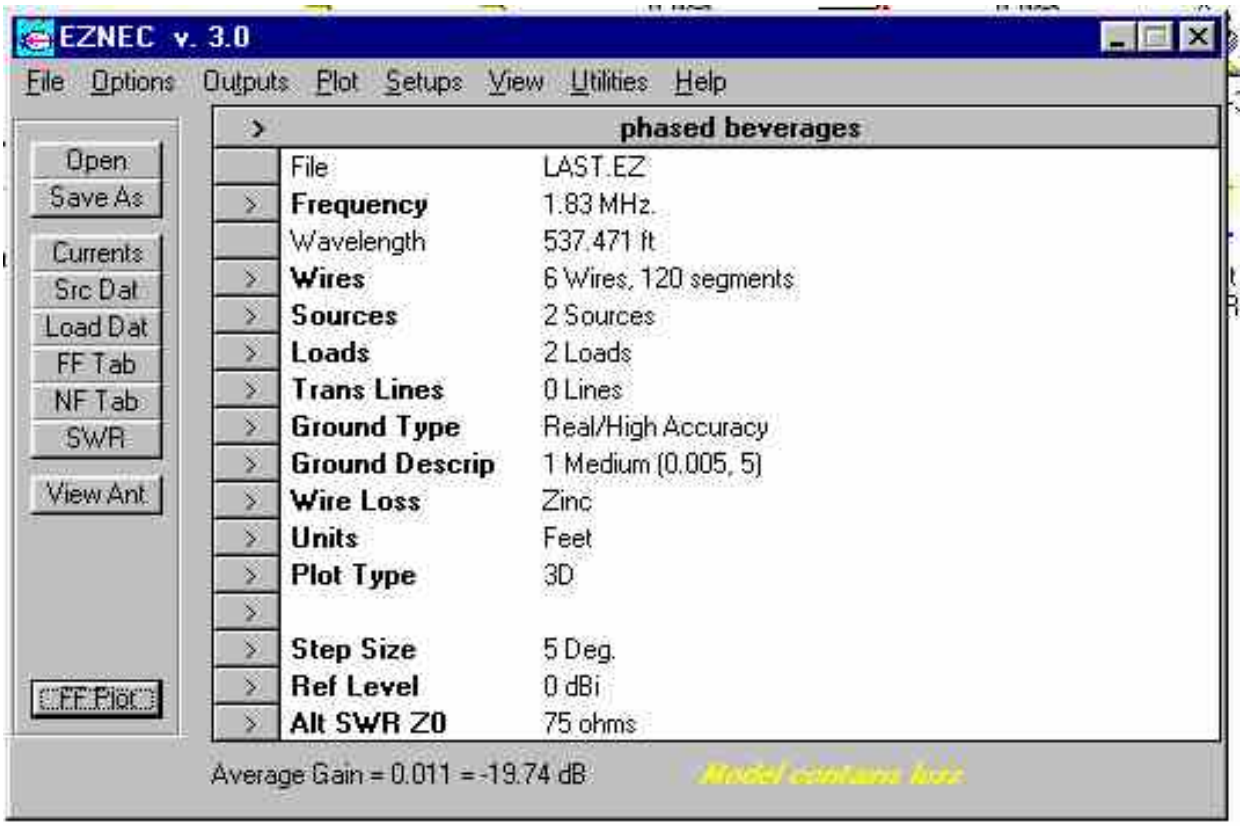
Gain is now -8.2dBi, about 3 dB better than the single wire's -11.24dBi. Did we gain any signal/noise ratio, for all the work of adding a second antenna? ***Not a bit...the S/N ratio is exactly the same because the directivity is the same.*** Since unwanted noise is external to the antenna, unless we have a change in pattern we can not have a change in S/N ratio. It is a common mistake, and absolutely incorrect, to compare gain and use gain to estimate receiving ability.

Eznec version 3.0 allows us to estimate directivity by modeling the antenna in three dimensions, and looking at the pattern. Here is the first Beverage modeled that way:



One source is set to 0 amperes, turning off one wire. Notice the "Average Gain" that appears in the lower area, -22.67 dB. If we subtract the average gain from the -11.24dBi gain at 20 degrees, we will have the ratio of response in the desired direction (from a two dimensional plot) to the overall gain of the antenna. This will be a positive number. The result is what we can consider a "receiving directivity factor" of the antenna expressed in dB. In this case, the RPF is 11.43dB.

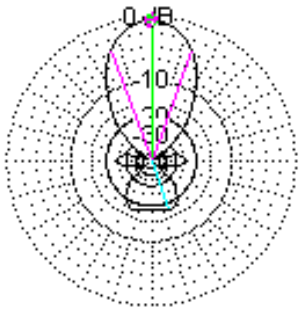
Now let's look at the two wire antenna, using the same method:



Notice I modeled this with both sources at 1 ampere, so both Beverages are active. Using the same process we used for the single source (which was the same as only having one Beverage) we see the RDF at 20 degrees is now (-8.2)-(-19.74) = 11.54db!

The effective receiving performance only increased 0.11dB, not worth the effort of adding a second antenna!

Changing spacing to 350 feet, we have the following two dimensional pattern:



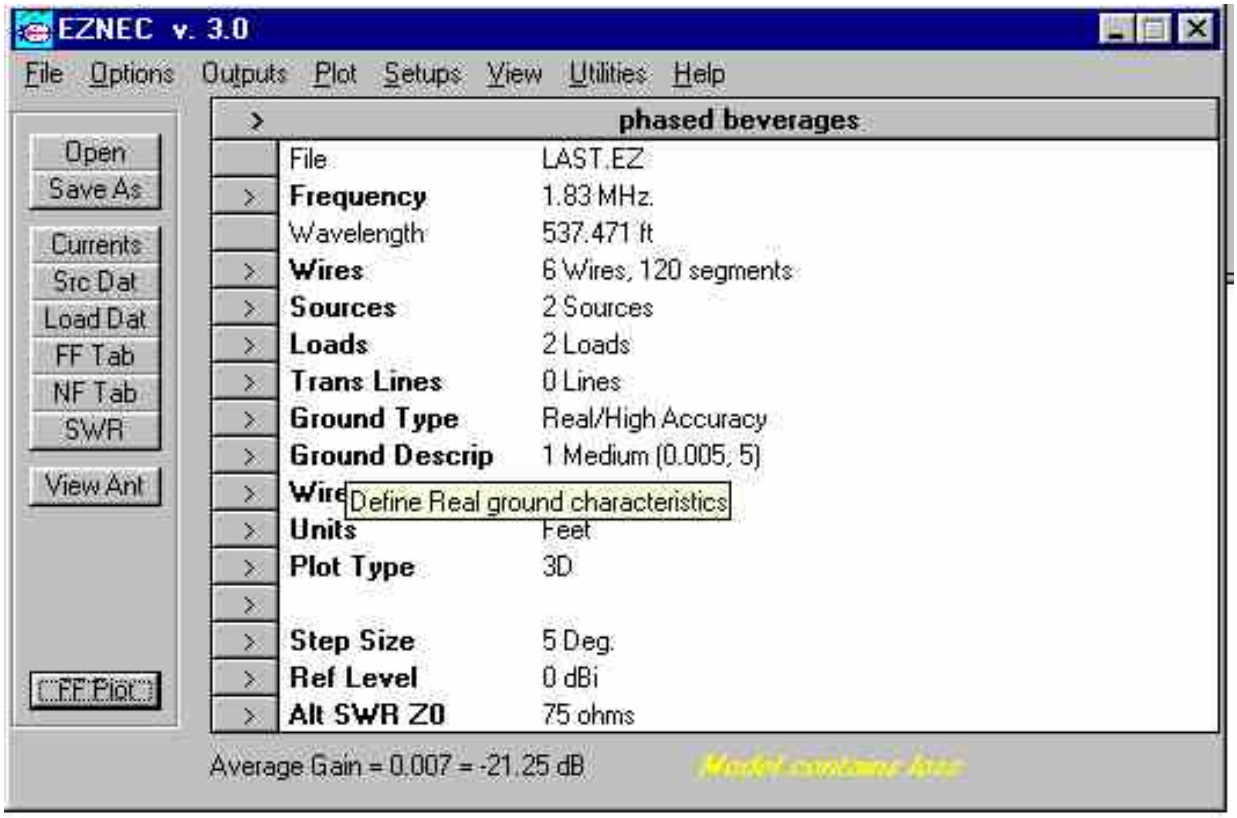
EZNEC

1.83 MHz

Azimuth Plot		Cursor Az	90.0 deg.
Elevation Angle	20.0 deg.	Gain	-7.86 dBi
Outer Ring	-7.44dBi		-0.42 dBmax

3D Max Gain	-7.44 dBi
Slice Max Gain	-7.86 dBi @ Az Angle = 90.0 deg.
Front/Back	17.83
Beamwidth	40.0 deg.; -3dB @ 70.0, 110.0 deg.
Sidelobe Gain	-25.1 dBi @ Az Angle = 290.0 deg.
Front/Sidelobe	17.24 dB

.....and the following data in the main menu of Eznec after modeling a three dimension plot.....



Taking (-7.44)-(-21.25) we now have a RDF of 13.81dB at a wave angle of 20 degrees. Comparing the three systems:

Antenna	Gain @ desired elev & azmth	Average gain	Rcvg Directivity Factor
single Beverage	-11.24dBi	-22.67dB	11.43dB
70 ft spaced Beverages	-8.2dBi	-19.74dB	11.54dB (+.11dB RDF)
350 ft spaced Beverages	-7.44dBi	-21.25dB	13.81dB (+2.38dB RDF)

Summary

Until Eznec's Windows version came along, it was very difficult to estimate performance. We often made mistakes, because we only compared gain. Using the above method, we can estimate antenna performance for evenly distributed noise. If there is no change in the Receiving Directivity Factor for similar pattern antennas, we can be sure we haven't changed overall performance for distant signal and noise sources.

There are a few things we have to keep in mind:

1.) If noise is not evenly distributed (which is often the case) performance will depend on the gain difference between desired signal direction (azimuth and elevation) to gain in the direction of noise(s). A 20dB null on the noise compared to gain in the desired signal direction will actually improve S/N ratio by 20 dB, if the noise from the null direction totally dominates all other noises.

2.) If noise arrives primarily from the same direction and angle as the desired signal (and assuming polarization of signals and noise is the same), there will be no S/N improvement.

3.) If noise originates in the near field of the antenna, all bets are off. Anything can happen.

All in all, the above method gives us a rough idea how various antennas compare in a environment with reasonably even noise distribution.

DX sound files

[Home]

Topband and DX Recordings and Sound Files

K1UO to JA on 160. The JA's are weak here!

These are mono and stereo recordings using either a FT1000D or phase-locked modified-R4C's with dual antenna systems for stereo diversity. This gives you an idea of how DX signals at this location under various conditions.

Clear band Beverage MONO directivity demonstration (W1AW-MONO-FT1000D). Every 5 seconds I switch directions. The first direction is NW, and I move around E, SE, S, SW, W, and NW. W1AW

Demo using storms in Texas and south. Every 5 seconds switch (right channel beverages, left channel verticals) NW, W, and SW. QRN

JA2 through noise, 10 seconds NW, five W, five SW. Receiver right channel phased Beverages, left channel Vertical Array QRN_ja2zjw

DX Stations On 160 Meters:

XU7ACB Long path (in evening) xu7acb121001_LP.wav

XU7ACB short path (my sunrise) xu7acb_am.wav

JT1/K4ZW short path (morning) jt1_k4zw112801.wav

VK9ML short path (morning) vk9ml041702_am.wav

BQ9P 160m (not on peak) Oct 11, 2003 around 1040Z

JA opening on good day Oct 12, 2003

Big file of Europe from east path during severe Aurora on Oct 30,2003

9V1GO from NW path (unusual path, 9V is normally skewed SW) Nov 03, 2003

On 80 meters:

A51B SW path 80-meters Feb 19 2003 1200-1215Z

Small Vertical Arrays

[Home] [Up] [Receiving 4-Square]

Small Vertical Arrays

While this is mainly intended as a guide to help install arrays, I have stock of all components in these antennas. If you are really stuck, I can provide turn-key systems that are tested and tuned, include phase-matched cables and phasing systems. I can supply custom arrays of up to eight directions using up to 16-elements from stock components and assemblies, tested on reliable commercial test equipment.

Never Pick Gain, F/B, or Take-off Angle as a Parameter for Weak Signal Receiving!

Arrays of small verticals provide excellent receiving performance when systems are designed and installed properly. As pointed out in other articles, there are key differences between receiving and transmitting systems. Parameters considered important in large transmitting systems are sometimes far from optimum in receiving applications, especially systems using small antennas. One key item is gain. Contrary to common opinion, more gain does not translate to better receiving, once the receiver is limited by external noise reaching the receiver. The key parameter is directivity, which may or may not have a parallel relationship with gain. Gain includes efficiency, directivity excludes efficiency. This important consideration applies to systems discussed below.

Noise rarely comes from one direction, or a narrow range of directions. With that in mind, antennas will be compared by a receiving directivity factor calculated with Eznec's (version 3) average gain. Removing efficiency from the equation allows direct comparison of receiving systems since directivity, not gain, is the determining factor in selecting an HF or LF receiving array. S/N ratio is very dependent on nulling or rejecting unwanted signals or noise.

It is important to locate receiving antennas as far as possible from radiators or re-radiators of unwanted signals and noise. Always remember noise has exactly the same characteristics, so far as an antenna is concerned, as signals from *intentional* transmitters. There is no way to sort "good signals" from "bad noise" except through the directional characteristics of your receiving antenna. Noise is not electric field dominant. Desired signals are not magnetic field dominant. The field impedances are all the same except near the antenna or source, and near the antenna or source coupling to multiple unknown sources is largely unpredictable. The truth is, it is anyone's guess what field impedance is actually best!

A small loop antenna, at a distance of a few meters, is magnetic field dominant. Here is an important fact few people, outside of those who work with nearfield systems know. At a distance of an eighth wave and larger a small magnetic loop becomes electric field dominant! Conversely, a small voltage probe becomes magnetic field dominant at about the same distance! The fields reverse dominance because of phase shift between the fields as the radiation fields start to overtake the induction fields.

Susceptibility to unwanted near-field and induction field coupling between receiving antennas and large transmitting antennas or noise sources is obviously largely unpredictable, although many problems can be corrected through changes in antenna placement or detuning structures and/or canceling the radiation from surrounding structures. When dealing with nulls, a modest amount of re-radiation from surrounding conductors can make a large difference in system performance, but the key is to watch overall directivity. Anything that reduces directivity will reduce the S/N ratio of a receiving system. The reduction is directly by the amount of null reduction ***ONLY*** when noise comes very predominantly from within the area encompassed by the deeper areas of null. ***We always, unless we have noise from a specific direction and angle all of the time, want a wider more modest depth null in favor of having a sharp point with a deep null and a wide-nosed response.***

One of my best arrays on Europe is only a few hundred feet from a transmitting four-square, clearly in the near field of the four-square. There is no detectable influence on this array when it "looks away" from the four-square, although there is a quite noticeable reduction of F/B ratio when beaming back into my transmitting antennas. The null to the SW is very deep, in excess of 35dB, regardless of four-square tuning. The converse is not true, the null NE when looking SW is only 10-15dB deep unless I detune the antennas. Yet the lower null depth barely causes a detectable noise increase, because the directivity does not change much. It is only when a very dominant noise (or QRM) arrives from the NE that this array becomes almost useless (compared to other arrays with deeper nulls in the NE direction). Of course detuning the transmitting antennas completely restores southwest performance, even though spacing is close.

Building Blocks For Arrays Using Verticals

There are four key areas overlooked in most published receiving arrays using small elements. Common oversights in element and phasing system design cause the antenna to be more critical to adjust, less stable, and provide a poorer pattern even if array elements happen to be working.

The most aspects are:

1.) Elements must be very low Q (wide bandwidth). They should have little reactance change with frequency or weather.

2.) Elements must be heavily swamped with loss. Mutual coupling effects must not change

element impedance.

3.) The phasing system must be designed for the impedances that actually appear at the phasing system. Transmitting-type boxes, because the systems have low loss, must have high mutual coupling effects and radically different element impedances. By definition, any given system *must* be seriously flawed in one application or the other!

4.) The phasing system must be stable and have very broad bandwidth characteristics.

Before building an array, we must select an element style that we can live with. Eznec and other programs have made this process simple.

The Basic Element

Length

A receiving array element should be as short as possible but still maintain sufficient sensitivity (gain) to ensure external noise exceeds receiver noise. My 160 Meter Band elements are about 20-foot vertical height. I've found all of my arrays with elements that height have overall sensitivities (gain) on par with my Beverage antennas, and that the signal levels are very easy to deal with.

Keep in mind that our systems require more gain as receiver selectivity is decreased. The noise floor drops in direct proportion to selectivity increase, and a change from 2.5 kHz selectivity to 250 Hz selectivity reduces noise voltage or power by 10dB. Signal level, however, remains the same for the same transmitted power within our receiver's bandwidth.

20-foot tall elements with reasonable element spacing always provide more than enough signal to operate through nearly 1/2 mile of F-11 CATV cable (similar to RG-11 or RG-8 cables in size) at my very quiet rural location ***before amplification.*** 3-5dB noise figure amplifiers are adequate to establish S/N ratio by arriving noise even when placed after the signals travel through those long cable lengths. To gauge my noise floor, a standard FT1000 with preamplifier "on" on my 200-foot vertical has less than S3 noise in the SSB position at mid-day. Unless you have less noon-time noise than that, you will certainly not require an amplifier at the antenna! The only exception is if you have very close element spacing, because close spacing decreases antenna sensitivity (gain).

Mechanical and Electrical Concerns

I use two basic mechanical configurations of elements. One system uses steel electrical conduit on 1/2-inch fiberglass rods (rods driven directly into the ground) with four "loading" wires, while my other system uses stronger chain link fence top-rail mounted on wooden posts. Both systems handled weather from ice storms (where the antennas were coated with almost a radial inch of ice) to high winds without problems. I have had no electrical connection problems, and no weather

detuning problems. ***It is not necessary to "insulate" the antennas mounted on wooden posts because system Q is very low and impedances are modest.*** Wet posts will have no deleterious effect on performance, although I would always place the loading system (and the base of the verticals) above snow depth. ***It is not necessary to use high-Q loading inductors.*** The only requirement is that inductors remain relatively stable in characteristics with climatic changes.

My systems are normalized at 75- ohms for several reasons:

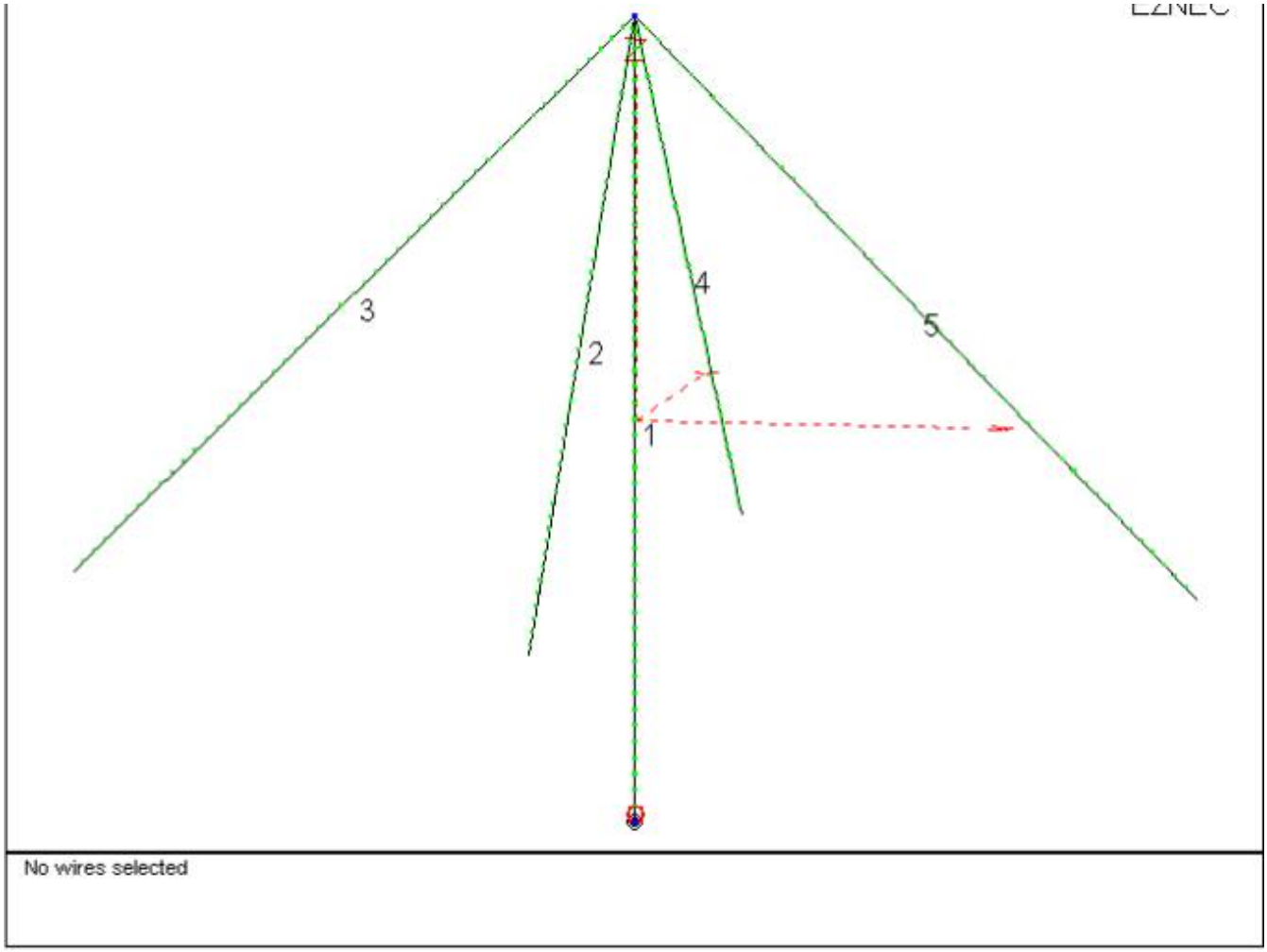
1.) 75-ohm feed produces a wider VSWR bandwidth than 50-ohm feed systems, the swamping resistive losses are about 50% higher.

2.) CATV cable suitable for direct burial is inexpensive and connectors are inexpensive, reliable, and easy to install.

I use F6 flooded CATV cable for local cables in arrays, and F11 (RG-11 size) or 5/8" flooded CATV cable for trunk leads. It is NOT necessary to use double or triple shielded cable, you gain nothing. But you do want to use good quality cable that will last years without weather changes.

Practical Verticals

My rectangular arrays use elements with four 20 foot long #16 loading wires, insulated by fishing line used to support the wires. I terminate the guying 20 feet out from the base of the antenna. The entire structure is self-resonant on 80 meters. The large "hat" makes current essentially uniform throughout the vertical element while minimizing unwanted sensitivity to high angle radiation, and also supports the elements.



Eznec Download

Eznec RXvrhat

This structure is base loaded with a series L/R combination of approximately 30uH and contains a ***total*** loss resistance of 75 ohms. This resistance includes resistive losses related to inductor Q, as well as ground system loss resistance. I used small molded choke inductors, although other components will work. My system requires only 56 ohms of lumped resistance to bring base resistance to 75 ohms.

Each element requires a ***stable*** ground system. Ground loss is not important, but long and short term loss stability with climatic changes is very important. I use a minimum of four buried radials, each 1/8- to 1/4-wl long, on each element. Always place radials directly under each hat wire. Do ***NOT*** use small elevated radial systems or grossly non-symmetrical radial systems! Elevated radials will reduce VSWR bandwidth of the array, introducing unwanted phase shift. They also make the system susceptible to high angle signals, and are more susceptible to common-mode noise on feedlines and other conductors around the antenna than buried or earthed radials. It is not necessary to bury radials, but if the radials aren't buried multiple ground rods are a good idea. The feedline should also be buried or if laid on the ground "choked" with high permeability ferrite beads near

each element.

Feedline Length

Unlike transmitting arrays, it is not necessary to use odd-quarter wave lines. It might be a tiny bit better if you use exact multiples of 1/4 wl, but even 1/2 wl lines work perfectly fine. Multiples of 1/4 wl work better in cases where you might fail to match antenna impedances to the transmission lines correctly. The reason of this is that phase shift in a transmission line is independent of line SWR when the feedline is ANY multiple of 90-degrees. With ANY phasing system having standing waves on the feedline, you can properly feed the system by supplying equal currents to any feedline an even multiple of 1/2 wl. Any feedline having odd multiples of 1/4 wl requires equal voltages feeding the line. The phasing systems I use, unlike transmitting systems, are designed to supply either equal voltages or equal currents! The proper ratio adjustments are easily made. You will, however, have slightly less phase error if you use any multiple of 1/4 wl when the lines are mismatched.

Beware that foam cables are NOT .82 or any other standard velocity factor. They range from the .70 range up to around .92 in velocity factor, depending on the ratio of gas to material in the dielectric. Only solid dielectric cable are predictable without measuring the cable.

Tuning the Elements

After careful planning and selecting the type of array, you should install the elements and the array's internal feedlines. Each element must be evaluated with an antenna analyzer that measures resistance and reactance. Connect the analyzer at the element's feedpoint, and follow these steps:

1.) Using a two foot long or less jumper cable, measure the antenna without a loading coil. Check the resonance for predicted values. It should be within several percent of the modeled self-resonant frequency.

2.) Touch the shield of the feedline in the array to the case of the analyzer or the vertical's ground. If impedance changes more than five percent, you need a better ground system.

3.) Install the loading inductance predicted, and sweep the desired frequency range for lowest SWR and zero reactance. If your analyzer is working correctly the lowest SWR will be at zero reactance, or very close to that frequency.

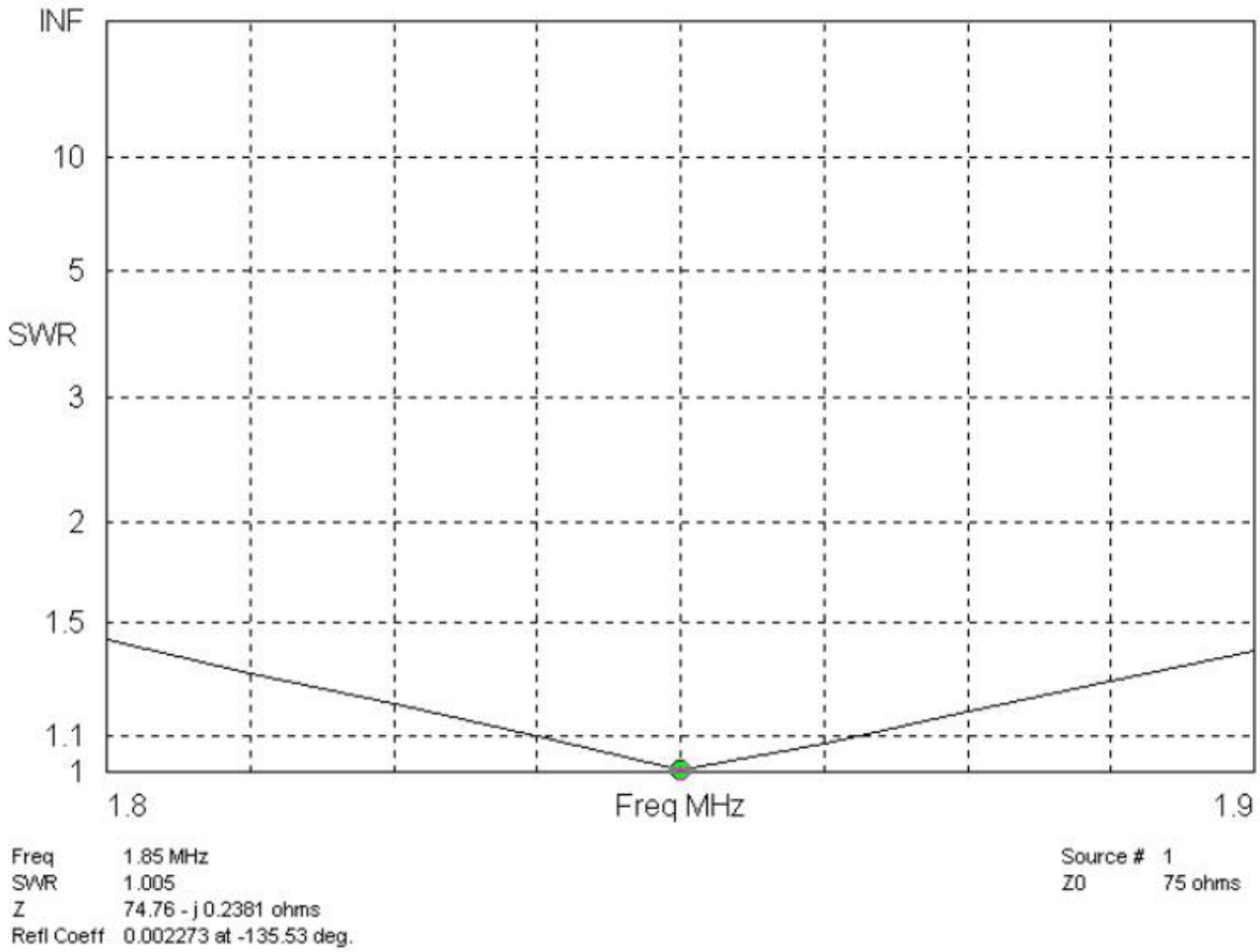
4.) Fine tune the inductance to make the antenna resonant at the desired frequency. You do this by adding or removing small inductors in series with a main inductor, or by adding or removing turns. Fine adjustments can be made by squeezing or spreading turns on the main loading coil's form, if it is a toroid or non-potted construction.

5.) Add enough series resistance to bring antenna impedance to 75 ohms at resonance (assuming you use 75 ohm cables).

6.) Check the feedpoint again for stability by connecting and disconnecting the shield of the unused array transmission line from the case of the analyzer or the ground system connection point. Again it should remain within 5%, or you need to improve the ground. If you can not improve the ground, you will have to isolate the ground by using a choke balun on the feedline.

Once one element is tested and proven, you should be able to duplicate that element with near-perfect results. I remove the matching system and take it to the test bench, and find a series C/R combination that produces the same resonant frequency. Multiple networks can be constructed on the test bench, and then moved to each element. My elements normally fall within 20 kHz of each other, any large difference in impedance or resonant frequency is a sign of potential performance problems.

Here is the 75 ohm SWR plot of this element:



Bandwidth is excellent, and sensitivity (gain) including all losses is -13 dBi. This element is almost perfect for use in small receiving arrays, since signal level and bandwidth are very good. Because of

the large amount of capacitance and the resistance loading, the element will not significantly change phase or sensitivity with frequency over the entire 160-meter band! In addition, it has very little response to high angle signals (and noise).

In circular arrays, hat wires can be extended to ~35 feet with only three loading wires used. My 350-foot diameter arrays position two loading wires (using tarred nylon fishing net string for insulation) in line with the perimeter of the antennas, while the third wire on each element is used to "pull out" from the array center. This tensions the perimeter wires and guys the entire structure. Even though hat wires are not spaced exactly 120 degrees apart, the effects on sensitivity to high angle signals are insignificant.

Arrays of Elements

Arrays of short verticals have both advantages and disadvantages compared to other antennas, such as Beverages and elongated loops. The comparison is:

	Elongated Loops	Beverages	Verticals and Wide-spaced Beverages
Area required	small <3/8-wl	large >1-wl	1/2- to 3-wl
Effort	modest	minimal	modest to large
Pattern	worse than 2-el vert, poorest	about same as 4-sq, middle	middle to best
Bandwidth	multiple bands	multiple bands	single/multiple band
Signal level	Low	Modest	Modest

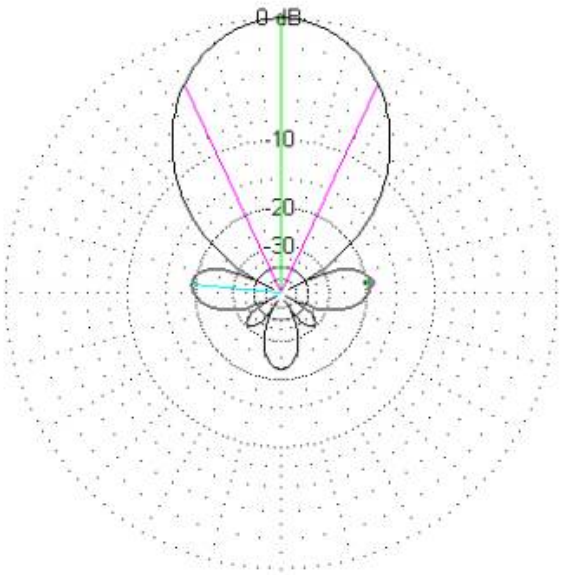
The primary advantage of arrays of short vertical arrays are excellent pattern and reasonable signal levels. Unlike balanced elongated loop systems, they are non-critical for feedline routing (other than keeping the feedline on the ground), matching transformers (transformers are not even required!), and earth conditions around the antenna. They have the same approximate output as simple Beverage arrays, and can still be made to work over very wide frequency ranges. The disadvantage is they are more complex, and require bandswitching to work on two bands.

Note: *I am finishing the layout of an FET amplified array that will allow ten-foot non-hat verticals that work from VLF to 80-meters and above, with the same basic directional characteristics over that range without switching.*

Broadside-Endfire Array (soon to be an

article)

The most directive four-element antenna possible is a broadside-endfire array. I use 70 feet endfire spacing and 330 feet broadside spacing between endfire cells, and get a pattern like this:



This array has an RDF of 13dB (two dB more than the large 4-square), and a HPBW of 47 degrees! The performance is similar to a circular array with eight elements in a 350 foot diameter circle, that allows directional selection every 45 degrees. The circular array uses techniques similar to the broadside-endfire, except the relay system is modified to switch eight antennas.

My next planned array is an array of four four-squares modified to fire eight directions, with 500 foot center point separation between each four-square array. I've found that 1000 feet is about the maximum physical separation allowing signals to be reliably combined on 160-meters. Arrays occupying areas larger than 2 wavelengths have far too much random phase and amplitude shift, preventing reliable combining of signals.

Receiving 4-Square

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Receiving 4-Square Array

Four small elements in a square layout, between 1/8- and 1/4-wl side length, can be used to form a four- direction array. Performance will be approximately equal to a 1 wl Beverage. The elements can be similar to those described in small vertical arrays.

As mentioned earlier, conventional transmitting phasing systems might be easy to obtain, but they are *NOT correct* when used in receiving arrays with short elements. Since each element has the same impedance in a lossy receiving array, the phasing unit "sees" 75-ohms from each element. When the antenna presents equal impedances at each port, the phasing system should source equal power at each port (which is also equal voltage and equal current)!

Use of a single 180-degree phase-inversion transformer for the center element along with delay lines approximately equal to effective element spacing create a very wide bandwidth phasing system. ***The phasing system is usable over at least one octave with appropriate element spacing and design!***

Upper and Lower Frequency Limits

Assuming you have ***broadband active elements*** of good design, the array will be useful when array side-length ranges from ~1/16wl spacing to 1/3 wl. The phasing system and delay lines are usable at any frequency, but the array spatial delay will provide useful end-fire patterns where array side-length is approximately 1/3-wl or longer. The appearance of grating lobes and dimpling of the front lobe limits upper frequency usefulness.

As frequency is decreased, ***array sensitivity*** eventually drops to unusable levels. The effect comes from two effects:

- Elements become electrically shorter, reducing element and array sensitivity.
- Element spacing becomes smaller in electrical degrees, reducing array sensitivity.

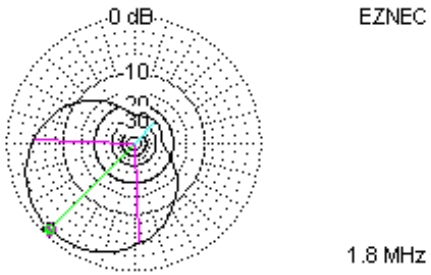
Shorter elements, in electrical degrees, mean less sensitivity. Sensitivity obviously must drop with frequency, for a fixed element height.

Element nulling always fully subtracts with reduced spacing. With any fixed spacing distance, a lower frequency array sensitivity limit appears when "forward" voltage phase vectors become too close to -180 degrees.

Element and array phasing-sensitivity eventually push the antenna into the receiver system's noise-floor, and this becomes the lower frequency cutoff. Pattern is maintained down to that cut-off, assuming the elements remain matched to the feedlines.

Even without broadband active elements, multiband or broadband arrays are practical with ONE fixed element group and a single unchanged phasing system. Element resonance and resistance would be the only parameter that would require switching.

Pattern with 70- foot per side spacing, on 1.8 MHz is:

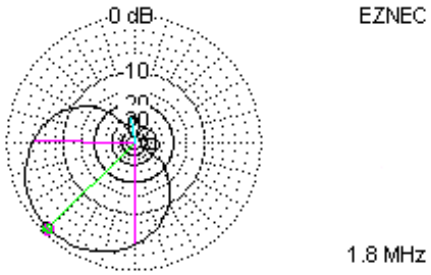


Azimuth Plot		Cursor Az	225.0 deg.
Elevation Angle	20.0 deg.	Gain	-11.51 dBi
Outer Ring	-10.59dBi		-0.92 dBmax

3D Max Gain	-10.59 dBi
Slice Max Gain	-11.51 dBi @ Az Angle = 225.0 deg.
Front/Back	23.19
Beamwidth	95.0 deg.; -3dB @ 177.5, 272.5 deg.
Sidelobe Gain	-34.7 dBi @ Az Angle = 45.0 deg.
Front/Sidelobe	23.19 dB

There is a slight distribution problem with smaller size arrays because of mutual coupling. This can be compensated if 1/4-wl feedlines are used on the lowest band with the phasing coupler shown. RDF (receiving directivity factor) is around 10dB. This is about 3dB higher than the best single Pennant antennas, and on par with a very long Beverage.

If the array is 140 foot per side, the pattern is:

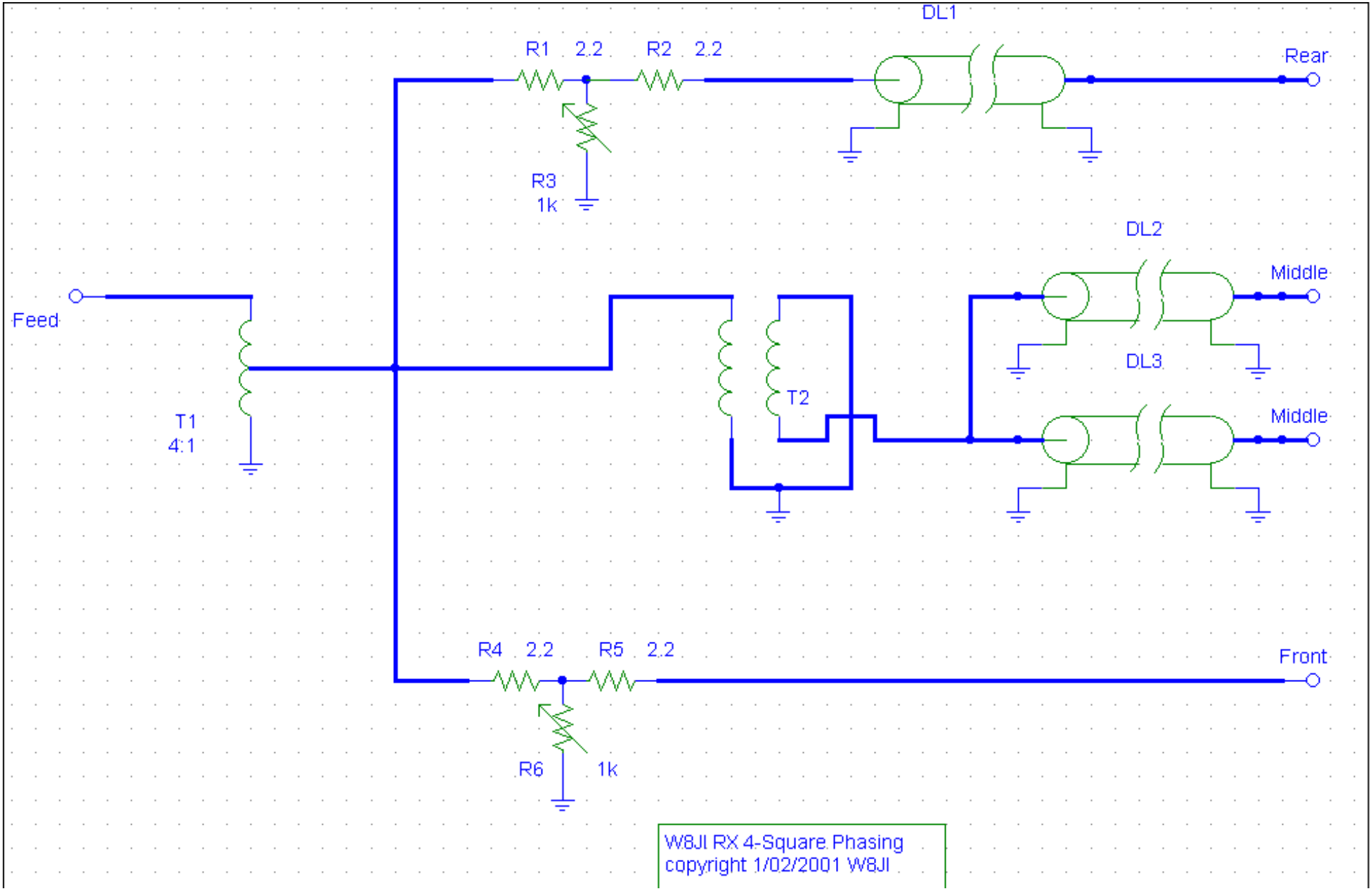


Azimuth Plot		Cursor Az	225.0 deg.
Elevation Angle	20.0 deg.	Gain	-6.88 dBi
Outer Ring	-6.14dBi		-0.74 dBmax

3D Max Gain	-6.14 dBi
Slice Max Gain	-6.88 dBi @ Az Angle = 225.0 deg.
Front/Back	31.17
Beamwidth	91.2 deg.; -3dB @ 179.4, 270.6 deg.
Sidelobe Gain	-33.16 dBi @ Az Angle = 95.0 deg.
Front/Sidelobe	26.28 dB

Doubling array size only provides about 1dB more directivity (note that HPBW only changes a few degrees), although signal level increases about 4dB! This array would make an ideal portable antenna for limited-space two-band operation, a SPDT relay at each antenna could be used to switch loading networks for each band.

The phasing system should look like this:



T1 is a 4:1 impedance ratio (2:1 turns ratio) auto transformer. A suitable 100 kHz to 20 MHz transformer would use 5 to 10 turns of #22 wire (wire size is not critical) in a twisted pair, connected with opposite ends in series. A Fair Rite Products 2873000202 core is recommended, and available from me if you can not find one. (A turn is one complete pass of the twisted wires through both holes in the binocular core). This transformer steps the 18.75-ohm impedance, produced by paralleling four 75-ohm lines, up to the feedline output impedance of 75-ohms.

T2 is a 1:1 ratio phase inverting transformer. Wire size is not critical as long as the winding fits the core. A good choice is a twisted pair of #22 wires, making five to ten passes through a Fair Rite Products 2873000202 core. This is the proper core! This is the same transformer used in T1, with a different circuit connection. Again, I will help you with complete tested transformers if you can not locate materials.

A low attenuation value ~ 75-ohm attenuator is inserted in the line to the front antenna and rear antennas. This attenuator compensates for slightly different losses in transformers and delay lines. Although it absolutely isn't a requirement when using wider spacings (1/4 wl) with proper transformer and element construction and materials, they can be used to "tweak" the array to extremely high F/B ratios. The attenuators become more and more important as the array becomes smaller or less perfectly constructed and tuned.

DL1 should be slightly shorter electrically than the (around .9 times the) corner-to-corner distance across the array (diagonal distance). This line MUST be the same impedance as the element feedlines.

DL2 and DL3 are exactly half the length of DL1. These lines must be the same impedance as the array's element feedlines.

NOTE: *The delay lines are slightly shorter than element spacing. This elevates the null a small amount above the horizon directly off the rear, forming a cone-shaped null reaching the ground an equal number of degrees either side of the rear. I generally use about 10-20 degrees less than the element physical separation distance of a line running between diagonal corners of the array for delay lines on the highest band.*

If you model this antenna, remember phase is inverted. Effective phase shift from the delay line actually becomes a phase lead that is 180- degrees around a circle from the delay you might expect. For example, a 90- degree delay line actually provides 90- degrees phase advance (or -270 delay) because of the inversion, while a 60- degree delay line provides 120- degrees advance (which is the same as 240-degree delay) .

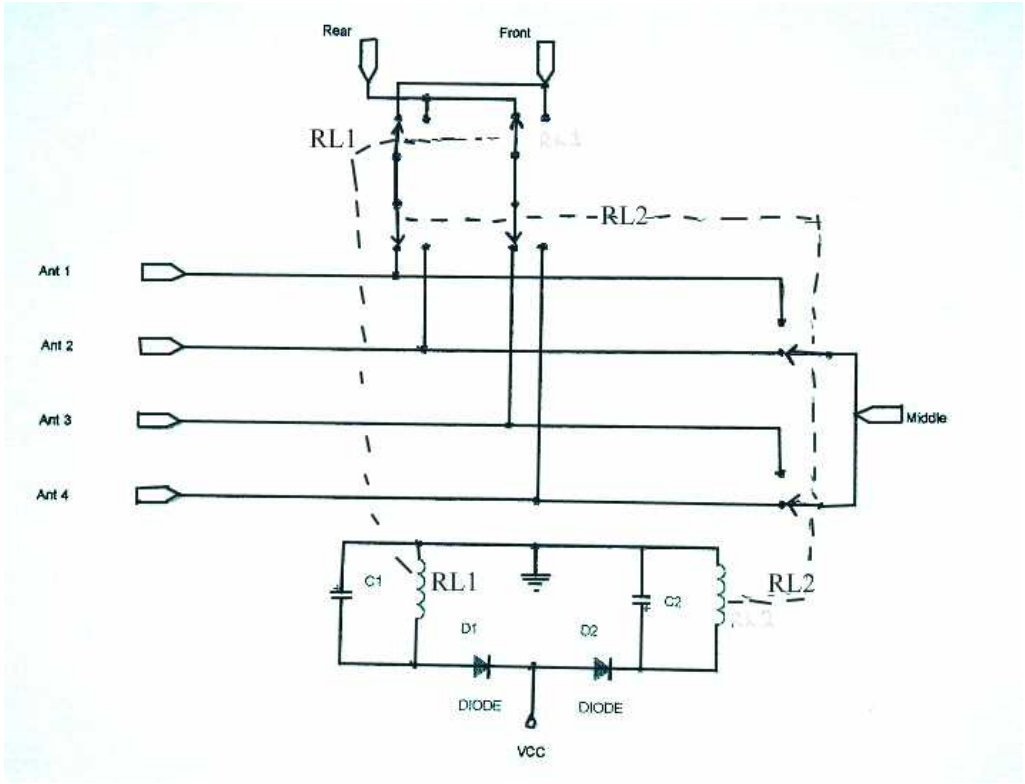
This "trick" causes phase-shift to track changes in element spacing with frequency, allowing the same delay line to work on multiple bands. The delay always is the proper length, which is slightly shorter than element spacing. It also causes the array to fire in the direction of the common-point element, since the feedline attachment point is the direction of lagging phase.

See the Crossfire phasing article.

Why Are the Lines nearly equal to the spacing for the elements they serve?

For an explanation, see the Cross-fire phasing page on this site!

The following system is useful for electrically rotating the array:



The antenna ports on the left from top to bottom start at the front of the antenna and from a top view go clockwise around the antenna. Zero voltage is position one at zero degrees, + is 90 degrees CW, - is 180 degrees, and AC is 270 degrees.

RL2 can be two separate DPDT relays with coils in parallel. I use 12V DIP relays with 220 uF 25v capacitors across relay coils. Any common silicon power rectifier will work for the diodes. If you have high RF levels near the control line, bypass it with a .1uF disc 50v capacitor. I route control signals through the feedline, but you must not have water in the feedline or poor shield connections to do that!

If you can not find parts, I have parts. I'm in the process of ordering commercial PC boards right now, to supply turn-key components kits or systems. I'd be happy to help with anything you need, as these projects move along.

W8JI_RX_ANTS

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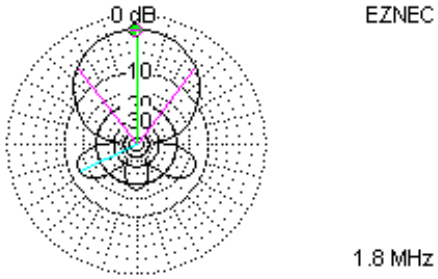
W8JI's Receiving Antennas

Conventional Beverage's ~580 feet long. Currently only two left, all others removed over the past two years due to lack of performance.

Specifications @ 15 degree wave angle:

Every 45 degrees from NE to NW. Directivity factor 7.29 dB, Half-power Beamwidth 78 degrees

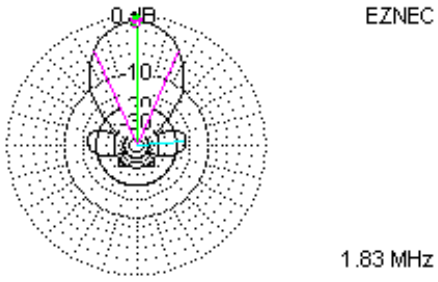


Arrays of Broadside Beverages 780 x 350 feet. Currently two duplicate systems (spaced ~1500 feet apart with antennas NE, E, SE, S, SW, W, and NW (28 individual Beverages make up this system):

Specifications @ 15 degree wave angle:

Europe, Africa, Asia/Japan, South America, ZL/South Pacific, VK/Pacific.

Directivity factor 13.8 dB, Half-power Beamwidth 42 degrees

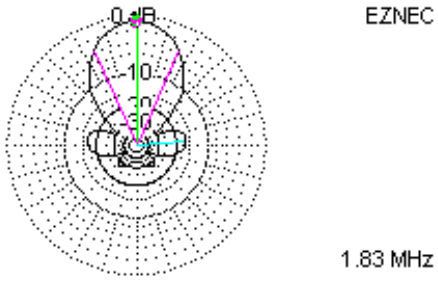


Arrays of Verticals 70 x 330 feet. One system:

Specifications @ 15 degree wave angle:

ZL/South Pacific, Europe.

Directivity factor 13 dB, Half-power Beamwidth 47 degrees



350-foot diameter circular eight-element array. One system:

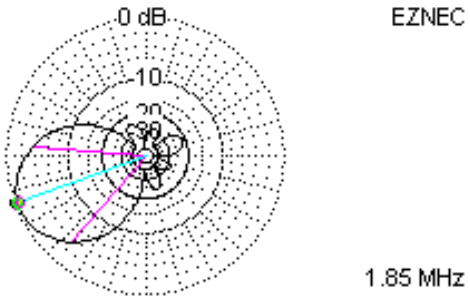
This array covers eight directions, spaced every 45 degrees around the compass. It has 22-foot tall top and base loaded elements. Over 1/4 mile of 75-ohm cable is used in this array (and 2500 feet of cable to reach the house).

Each antenna element is made from 1-1/2 inch chain link fence top rail. The fence rail hinges on these wooden posts with a bolt through the bottom of the vertical rail. The top of the wooden post has a steel pin that fits in a hole in the metal railing, and it is lashed in place.

Specifications @ 15 degree wave angle:

All compass directions. Electrically rotatable.

Directivity factor 12.1 dB, Half-power Beamwidth 53 degrees



RADIATED and CONDUCTED NOISE

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Switching mode power supplies, light dimmers, computer networking systems, poor connections that arc, and other "accidental transmitters" that either switch or spark can create considerable RF energy on wiring. It is helpful, when attempting to reduce such noise, that we understand how the noise travels from the source into the receiving system.

Definition of a Source

Let's consider the problematic device a "thing in a black-box", and call it the "***source"***. It doesn't matter if it is a computer, the spark in an electric fence, a light dimmer, or a doorbell transformer. The smallest area we can isolate creating the unwanted energy will be the ***source***.

Radiation from a Source

There is very little radiation from a source. As a matter of fact, even a very powerful single-terminal source can't transmit at any distance. In order to be an effective radiator, the source must have two or more connections to the outside world. The multiple connections (two or more) could be the two (or three) wires in a power cord, or a insertion point in a single wire. It could also be a connection to two totally different sources. Let's look at examples of each.

Arcing Splice

Consider a poor slice in a conductor carrying more than a few volts can cause a RFI generating arc. Two of the most common examples are an electric fence or power line.

The noise results when an insulated area with an electrical open circuit at low voltages breaks down as sufficient voltage appears across the gap. This ionizes air in the gap, and the resulting plasma conducts heavily. Such arcs can be particularly noisy at radio frequencies, because resistance can fall faster than voltage increases creating a negative-resistance. Negative resistances created in spark plasma were actually used as the amplifying element in crude oscillators. The plasma excited resonant circuits in the very first "single-frequency" transmitters, and at times those transmitters spanned distances of thousands of miles to very poor receivers using what were very poor antennas by today's standards!

Each conductor direction leaving the arc forms half of a large antenna. The entire system, from the source out in both directions, radiates. There isn't any effective way to cure or substantially reduce the unwanted radiation except by stopping the arc. Poor slices can become quieter in heavy rain, and can be broken up by wind or wire movement.

This type of arc excites the conductor entering the splice from one direction as one terminal, and the wire exiting the splice as the other terminal! The coupling mechanism is like a dipole feedpoint, and is very efficient at radiating over very wide frequency ranges.

Arcing to Another Conductor

A poor insulator or a close-spaced gap can break down and generate noise.

Once again, the noise results when an insulated area with an electrical open circuit at lower voltages breaks down when sufficient voltage appears across the gap. This ionizes air in the gap, and the resulting plasma conducts heavily. Such arcs can be particularly noisy at higher radio frequencies, because resistance can fall faster than voltage increases creating a negative-resistance.

Shunting-arcs generally have poor low-frequency energy, because the gaps are almost always connected to a very high impedance at low frequencies. This makes a very poor antenna connection for the arc, and reduces low-frequency response. If the gap had a path to a low resistance, the plasma would form a dead-short in the path. This would quickly cause a catastrophic system failure in systems able to supply sufficient current.

Non-arcing RF Generators

Many devices contain high-speed switching systems. These devices include but are not limited to televisions and monitors, computers, lighting systems with dimmers, and low-voltage lighting systems using switch-mode power supplies.

These non-arcing sources are similar to arcing sources, in that they require at least two terminal paths to become effective radiators. They generally are frequency selective or frequency periodic, and produce a broad buzzing signal that drifts around.

Common Characteristics

No matter what source is at work, we can be sure more than one conductor is at work in coupling from the source. If we disturb that path by isolating one conductor with a high RF impedance, we can reduce the interference. But by far the most effective way to remove or reduce interference, other than by removing the actual offending source, is by "shorting" the RF path with a bypass capacitor. As we will see later, this is almost always much more effective than adding series

impedance.

Common Noise Sources

Arcing Powerline Insulator or Hardware

The bell-insulator in a power line is probably the most common noise source. Older bell insulators has two interlocking but well-insulated metal posts. These metal posts have considerable area, because of the required mechanical strength. This results in considerable capacitance, in the order of a few dozen picofarads or greater. The ends of each bell terminal are held by a loose-fitting pin assembly. This pin assembly is the root of most powerline noise problems.

Bell insulators are often misused by power line crews. They are sometimes used to terminate short spans between poles, where the correct insulator would be a rigid post type or a polymer insulated fiberglass rod. The bell insulator, used in a short span, often is installed without sufficient tension. The lack of tension allows corrosion to set-up on the loose-fitting pins, and the capacitance of the insulator creates a voltage divider. Since the pin capacitive reactance is very high compared to the insulators leakage resistance and capacitance, a very large voltage can appear across the pin. This is true even when the pin is connected to grounded hardware! These loose short spans are called "slack spans", and bells should NEVER be used on slack spans! Slack spans with bell insulators generally become quieter during damp weather or in or just after rain, and can often be broken up by wire movement.

As the sine-wave peaks, the corrosion or oxides in the joint of the loose-fitting pin break down and arc. This causes an unintentional electrical noise to be generated, and the two terminals for the transmitter become the power line conductor and the grounded pole or hardware on the pole. With a wooden pole, the entire upper area of the can be excited with leakage currents from leakage capacitance and leakage resistances. Even if an insulator isn't arcing, a loose metal staple over a grounded wire or other poorly connected metal-to-metal joints will arc. Even a very tiny arc can excite the ground wire running down the pole and the power line wires as two (or more) terminals for the source.

To a much lesser extent transformers, lightning arrestors, fuse assemblies, and disconnects can have internal arcs.

Trees and foliage may also contact HV power lines and "burn" against the wire. Generally foliage noises are not extremely loud sources at low frequencies because one terminal of the source is through the tree, but occasionally they can be strong at lower frequencies.

As a general rule, but not always:

- Series-connection arcs from loose or corroded metallic hardware are reduced by wet weather. Water, contaminated by the corroded metals, forms a path around the poor connection and stops or reduces the arc.
- Shunt connection arcs from poor insulation are often reduced and may disappear in dry weather. They are caused by dielectric (insulation) failures.
- Corona noises are stronger at very high radio frequencies
- Hardware arcs and bells are generally strongest at higher radio frequencies
- Hard arcs like splices, clamps, transformer insulation, and switches are generally worse at lower radio frequencies
- The arc can be a very long distance from the point where you hear it on LOWER radio frequencies

Electric Fences

Older electric fences had a few-second timed off-and-on buzz. Newer fences have a "popping" noise, because they charge the fence in "tics". Fence problems are similar to power line problems, except of course the hardware is much smaller! Most severe problems are related to poor insulators, bad splices, or loose hardware.

Computers, TV's, and Switching Systems

In older times, TV sets created problems from sweep circuits. The horizontal sweep circuit operates with a saw tooth shaped waveform that was rich in harmonics. Early video systems overpowered the horizontal sweep system, and used the extra energy to supply high tension for the CRT. This resulted in many point-to-point wires carrying high power harmonic-rich 15kHz RF energy. A typical 25-inch television would use about 15 watts of the horizontal sweep system power for sweep, and perhaps 30 watts for powering the CRT second anode!

Wasted energy from the 50-75 watt sweep system was often coupled to the outside world a push-pull signal between the antenna leads and the power cord.

Modern computers and TV's mostly create noise-problems from internal switching power supplies, rather than sweep systems. The sweep systems, if ever even required, now generally only power the sweep. The HV often comes from separate HV power supplies.

Switch mode power supplies, whether in a TV, computer, telephone, video, or lighting system are rapidly becoming the most common source of modern RFI problems. Most of this has to do with poor testing techniques, lack of good standards, poor enforcement of the few poor standards we have, and a lack of skill or knowledge in power supply and device design and installation.

Fortunately most problems can be corrected by external filtering, but it generally needs to be done at the offending device.

Curing RF Egress or Ingress

Once the outside world is reached, the push-pull nature of the source can appear as a common-mode signal following a group of wires for many miles. We can think of this as one very long wire with the earth as the return path. The signal either directly radiates into our antennas, or it excites our antennas via common mode paths along cable shields.

While power lines and high-tension electric fences must be cured by removing the arc at the source, lower voltage systems can almost always be cured through proper external bypassing.

As a first response, we often like to throw a few ferrite beads at the problem. In more sophisticated approach, we might use multiple-turn series inductors. What these approaches miss is the great difficulty in obtaining adequate series impedance. When we consider the series impedance has to be totally *ISOLATED* for each conductor, and further consider core saturation from low-frequency operating currents we quickly realize series impedance is ***NOT*** the most effective method. It is helpful and may reduce RFI to acceptable levels in mild cases, but it certainly is an ineffective cure. Even if we are successful at one frequency range, it becomes virtually impossible to have a high series impedance over very wide frequency ranges.

The best approach is to add some RF series impedance and to augment it with excellent bypassing of all leads entering or leaving a device to one common point. That common point generally is connected to a safety ground, available in the USA at the third ground terminal of every modern-code electrical outlet.

Coaxial or audio cables should all leave the device through a bulkhead plate with all shields grounded. The power cord should pass through that bulkhead, and be bypassed to the bulkhead with properly selected UL/CSA/VDE approved line-bypass capacitors. Any chokes should be installed between the source device and the bulkhead. When this happens, the device becomes an isolated entity all by itself. It can not excite outside world conductors with unwanted RF.

It is indeed fortunate that the very same things that cure RFI often lead to greatly improved lightning protection.

Pre-amplifiers

[Home] [Up] [Preamp Summary]

For the summary page click here pre amp_summary

The following is a comparison of HF low noise receiving preamplifiers. More amplifiers will be added as they become available to test. All instruments used for these measurements are currently calibrated newer-production laboratory grade instruments.

Detailed graphs will only be added if specifically required. The .jpg pictures on this page are examples of how measurements are displayed on the instruments used in tests.

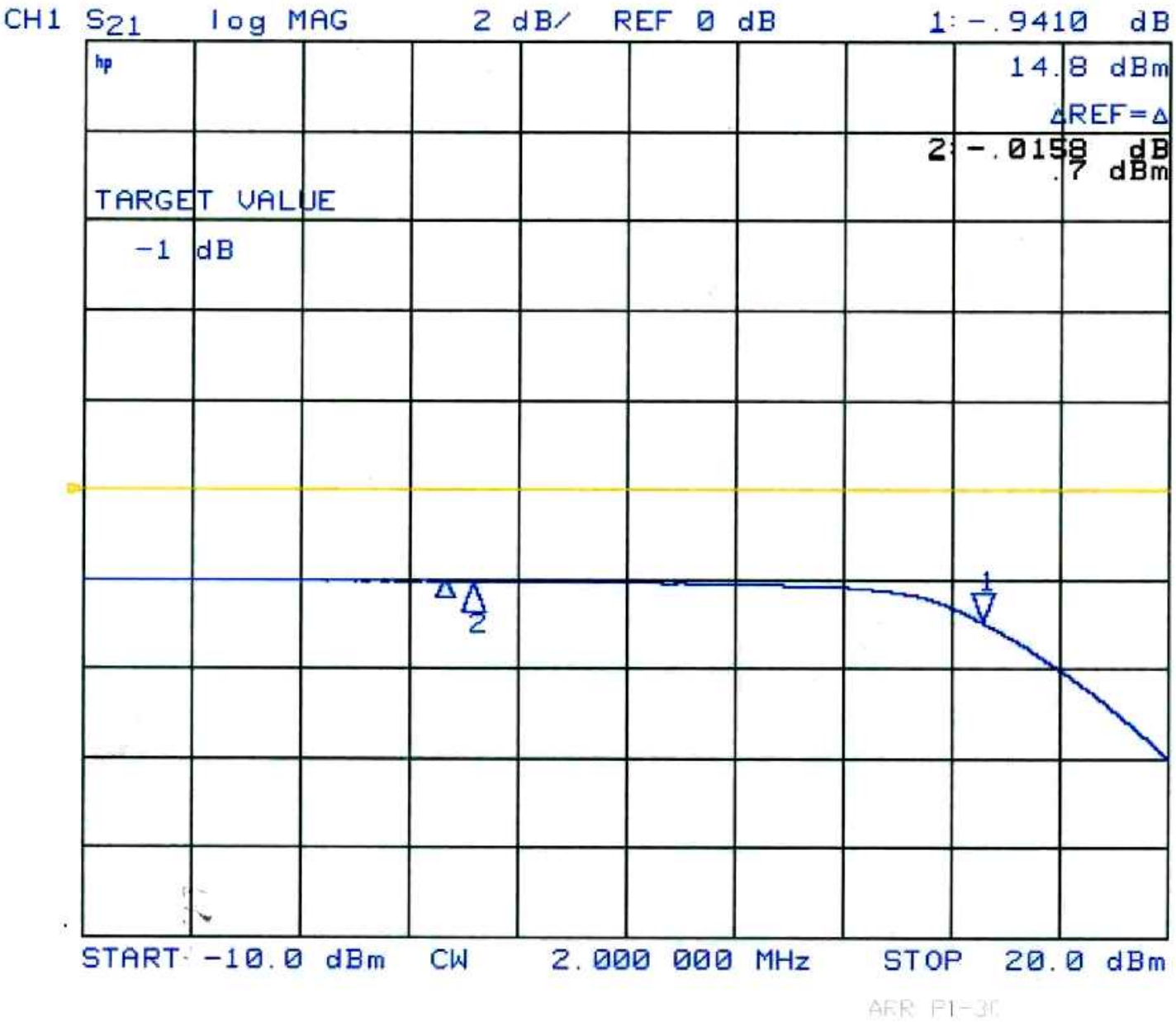
Gain Compression Tests

Pre-amplifier gain compression is measured on an HP (Agilent) Vector Network Analyzer and S-parameter test set using the power level sweep function at 2 MHz. A gain compression test shows loss of gain as input power is increased, or the input level where the amplifier starts to saturate.

This test indicates how much overall signal power the amplifier will handle before gain is reduced by non-linearity or saturation. This is power or gain non-linearity, and really does not directly represent distortion level vs. input or output signal power. A correction is applied to results for amplifier gain and equipment protection attenuators.

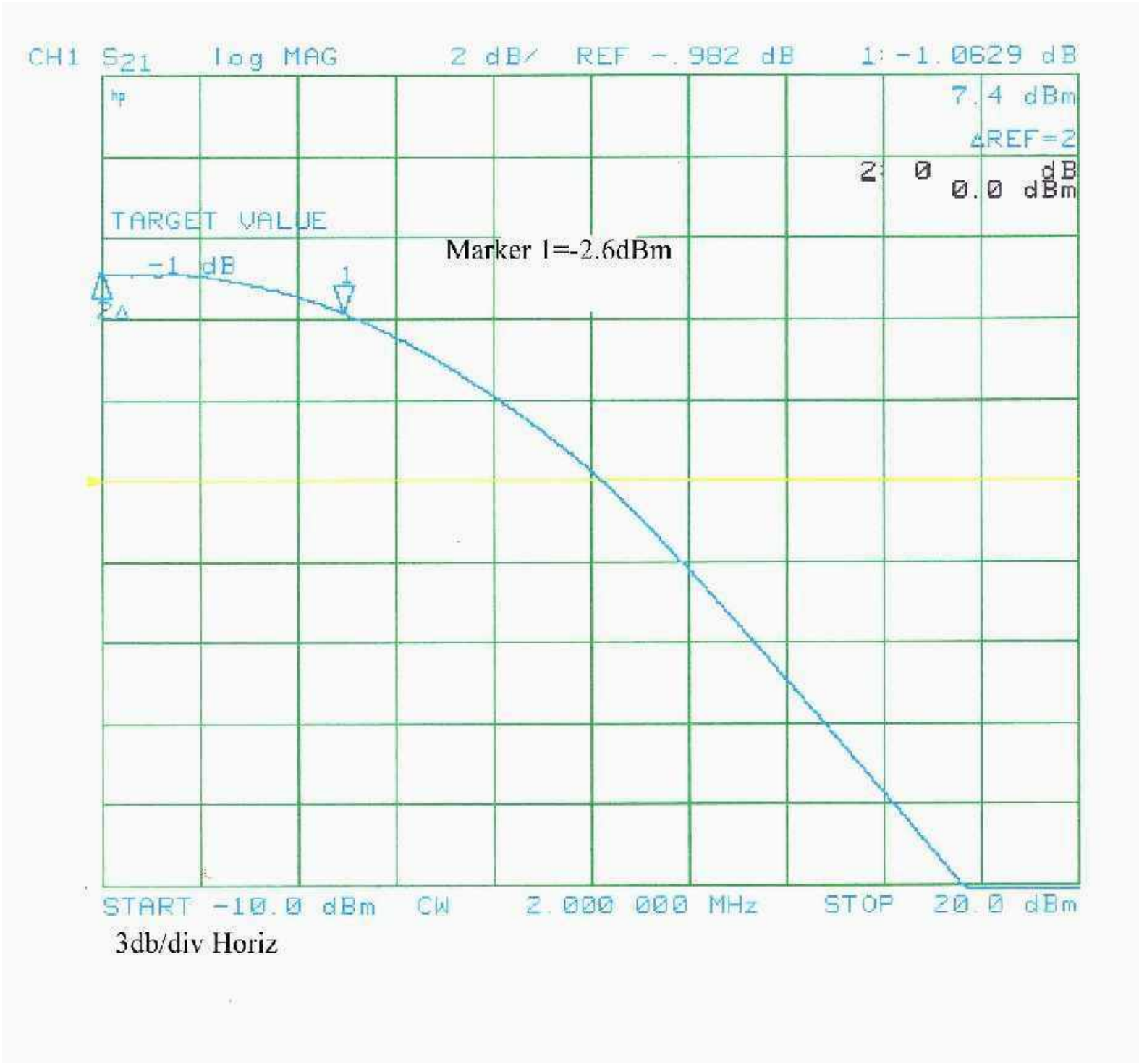
Note: *Third-order intercept is a better demonstration of signal level capacity. Third-order intercept better indicates immunity of an amplifier to producing unwanted signals (mixing products) from multiple strong signals.*

ARR 1-30



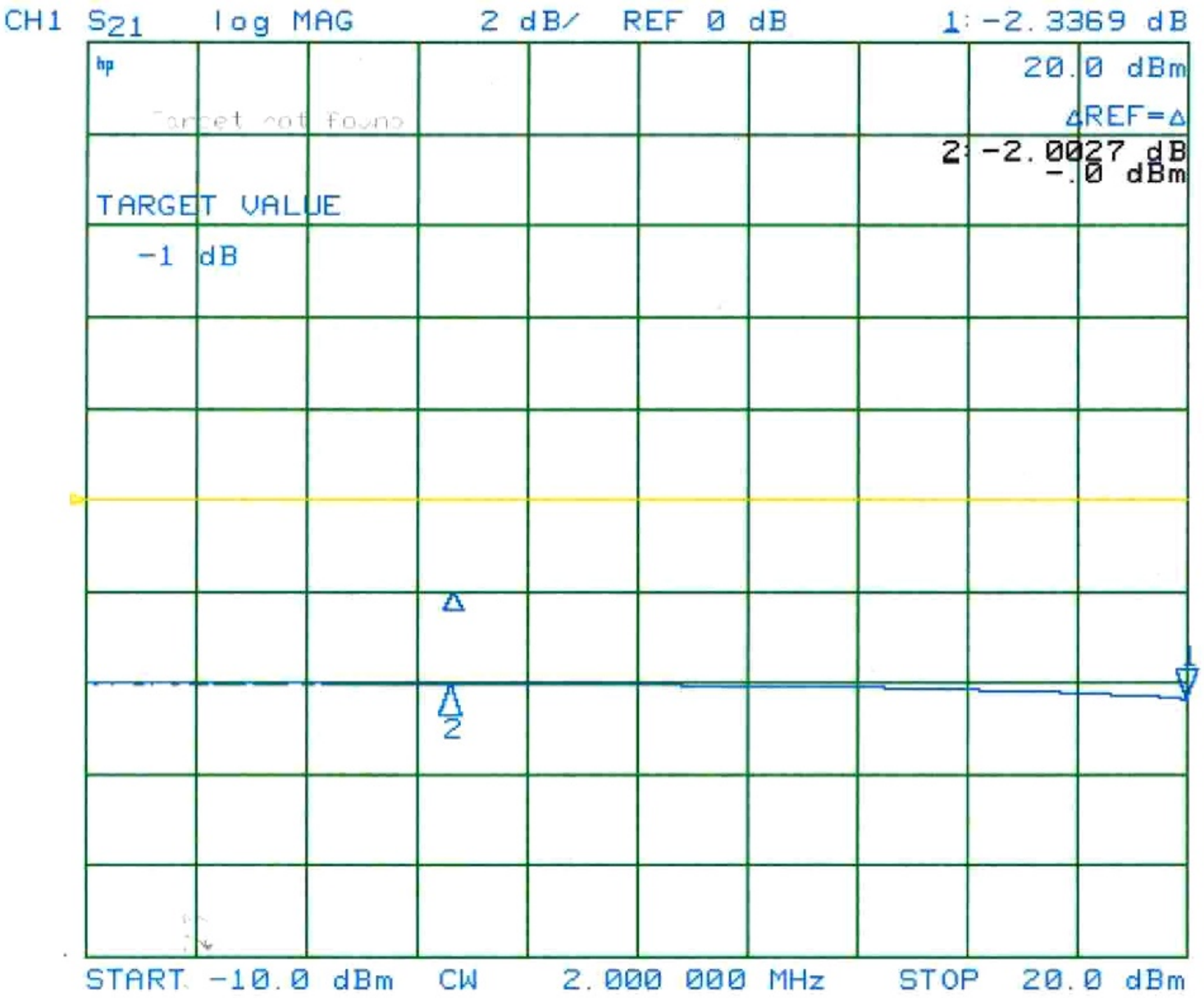
Amplifier input power (horizontal sweep) is 3dB per division. Measured 1dB compression is corrected to **21.5dBm output power @1dB compression**.

ARR GAsFET (1.8-2 MHz model)



1dB compression is corrected to 8.1dBm output power.

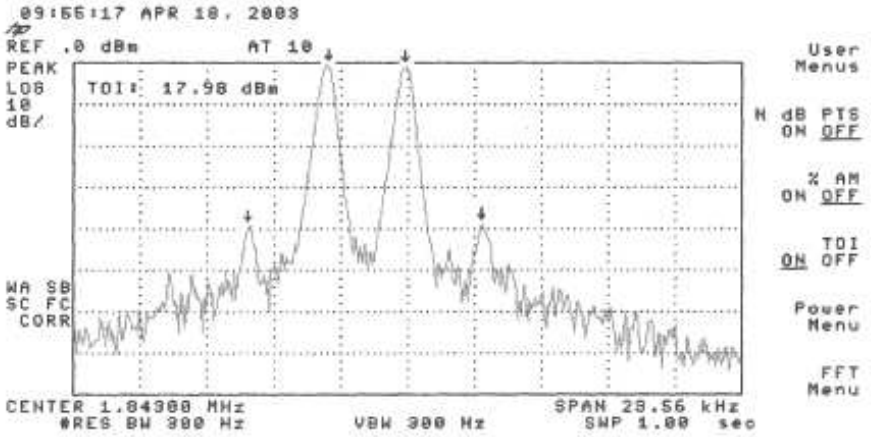
DX Engineering Push-pull HF preamplifier



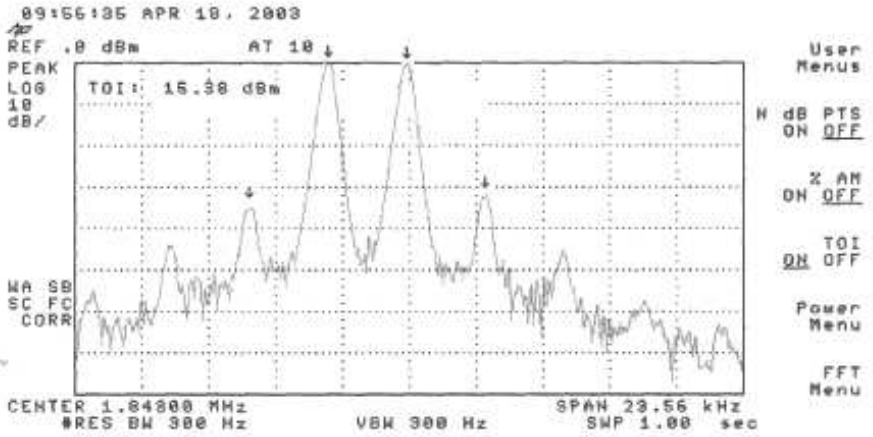
Gain compression beyond power output limit of analyzer. In other words, my test equipment does not have enough output power to drive the RPA-1 preamplifier into compression. Actual corrected gain compression (manual test) is approximately +26dBm (400-milliwatts output).

Comparison of Third Order Intercept

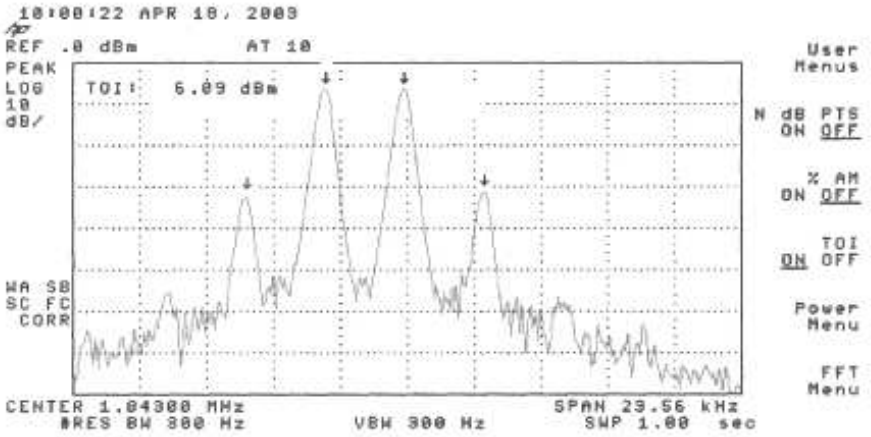
Third-order intercept tells us the immunity of the amplifier to generating spurious signals in the presence of multiple strong signals. This test was at 1.85MHz using the automatic IM3 test measurement. A stepped attenuator pad at the amplifier output was initially used to prevent analyzer damage during testing. Because of this "protection" attenuator, actual TOI must be increased by adding the attenuator value to the analyzer display reading.



DXE Preamp



ARR1-30



ARR-GAsFET

Third-order IM3 levels occur at the following single tone (of an equal two-tone test):

DXE (early model) -0dBm input power (instrument TOI 17.98+20.2dB pad= 38.18dB TOI)

Note: Later model RPA-1 measured 43dB third order intercept

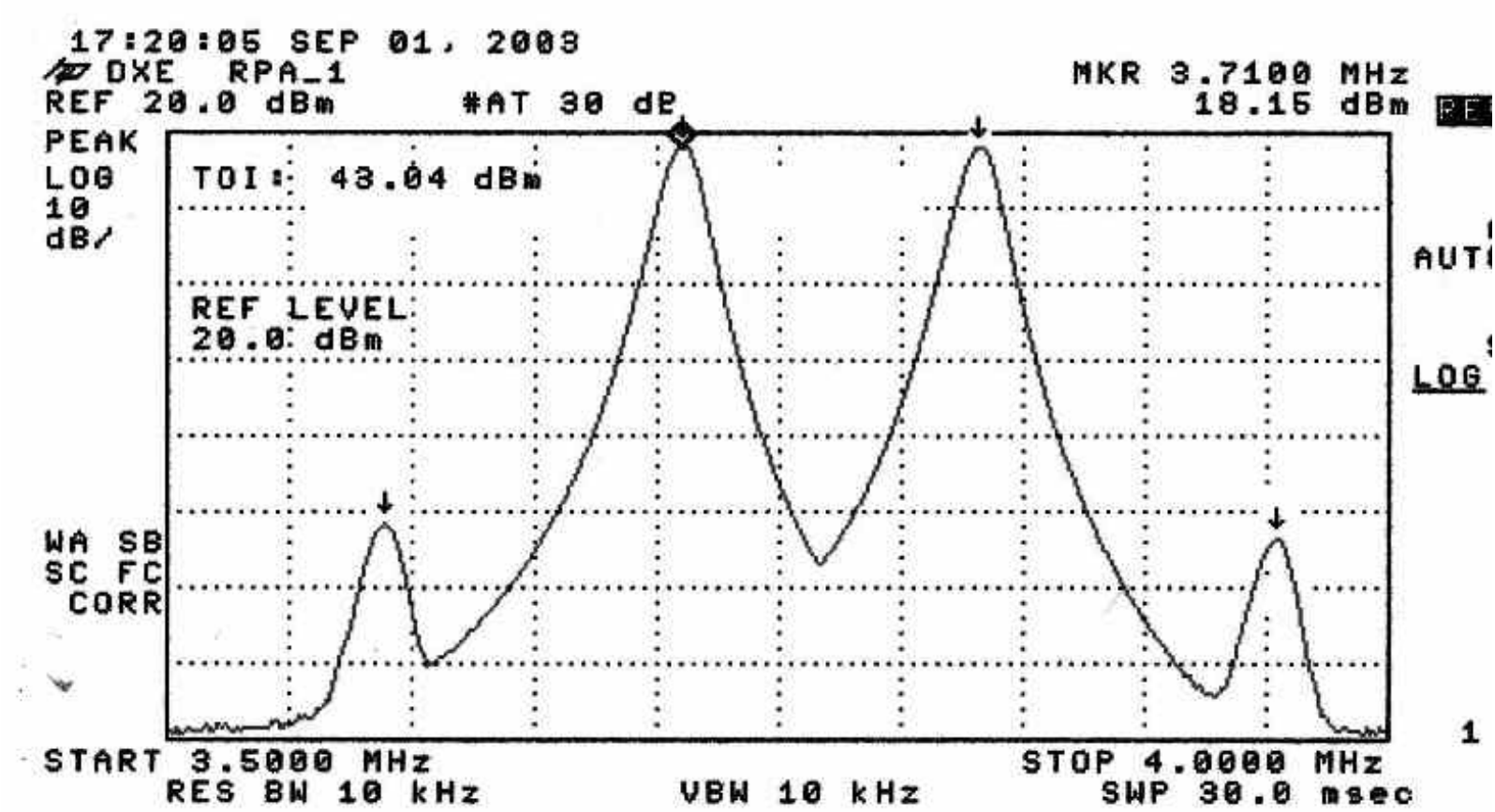
ARR P1-30 -5dBm input power (instrument TOI 15.38dB+15dB pad= 30.38dB TOI)

ARR GAsFET -16dBm input power (instrument TOI 6.09dB+5.3dB pad= 11.39dB TOI)

If you look at the above measurements, you will see third-order intercept in the DX Engineering amplifier occurs almost 8dB higher than in the better performing of the two ARR amplifiers that were tested. Not only is the DXE 3rd order intercept higher, higher order products are substantially less than the other two amplifiers produced.

DXE-RPA-1 Test

This is a later model DXE preamplifier model RPA-1. This measurement has no external attenuator, TOI in this case is a direct reading.



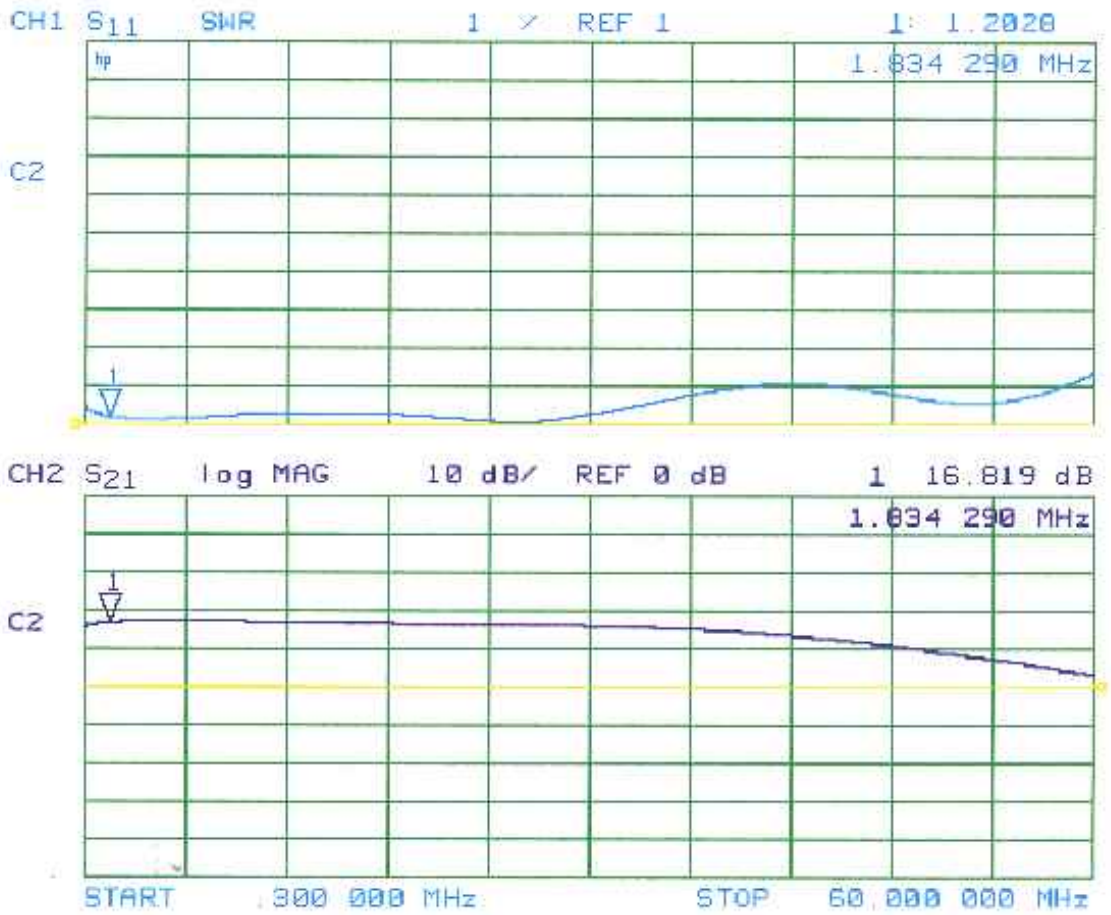
Preamp Noise Figure

I don't have this data saved to a file yet, but the noise figure of the GAsFET amplifier and the DXE amplifier are about the same. The ARR 1-30 had the highest noise figure, about 5.5dB. All of these amplifiers are acceptable for HF receiving in most applications so far as noise figure is concerned.

Input Impedance and Gain

So far, the DXE pre-amplifier has the closest match to 50-ohms over the range of 300kHz to 30 MHz. Gain is also reasonably flat in the DXE amplifier.

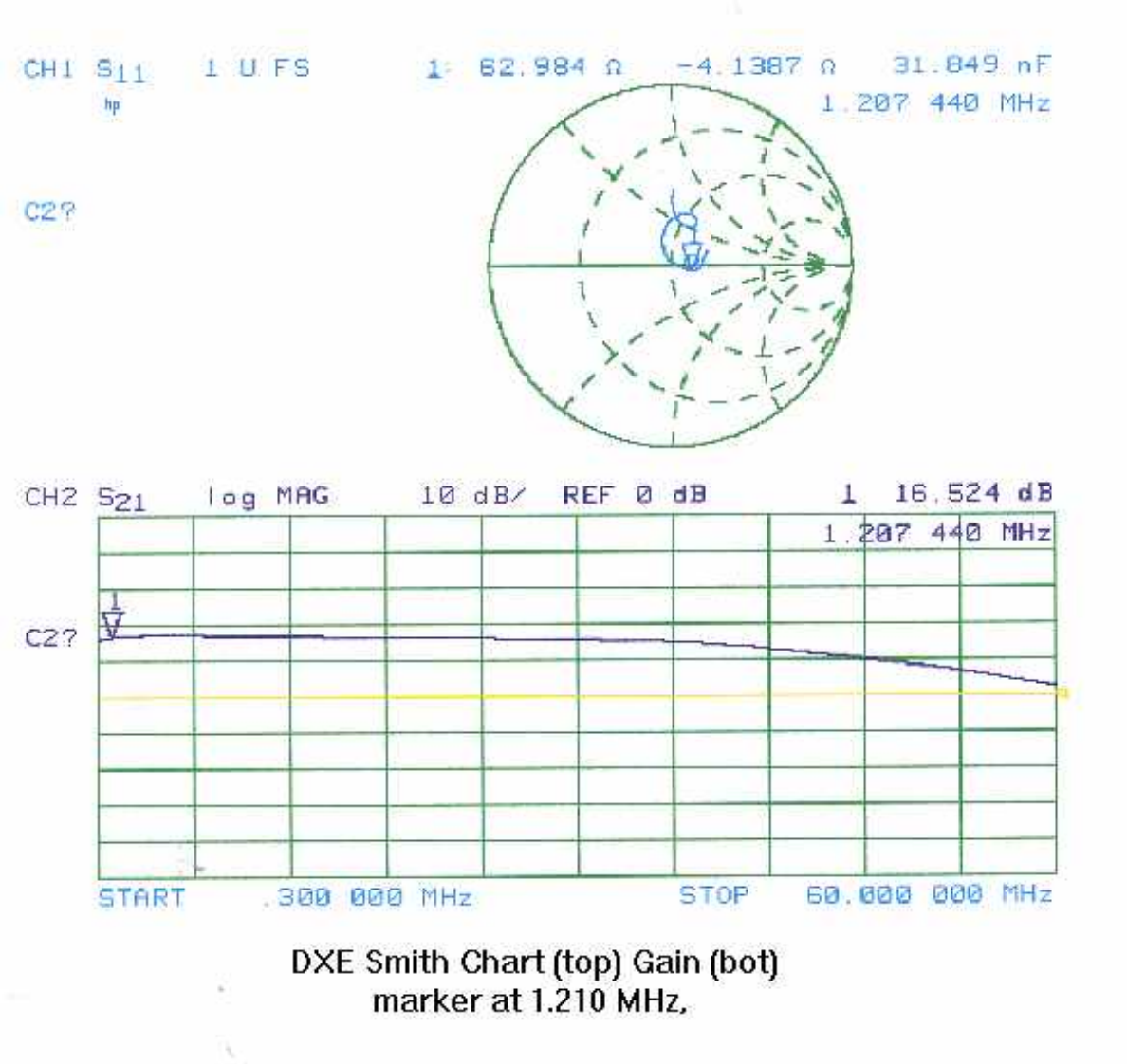
DX Engineering Preamp



DXE preamplifier SWR (top) and gain vs. frequency (bot)

The display above shows input SWR of the amplifier on the top graph, and gain on the lower graph. The marker is at 1.834290 MHz. The vertical graph lines are frequency at about 6 MHz per division. Gain at marker frequency is indicated as 16.819dB, SWR is 1.2028:1.

It is also possible to display input impedance on a Smith Chart:

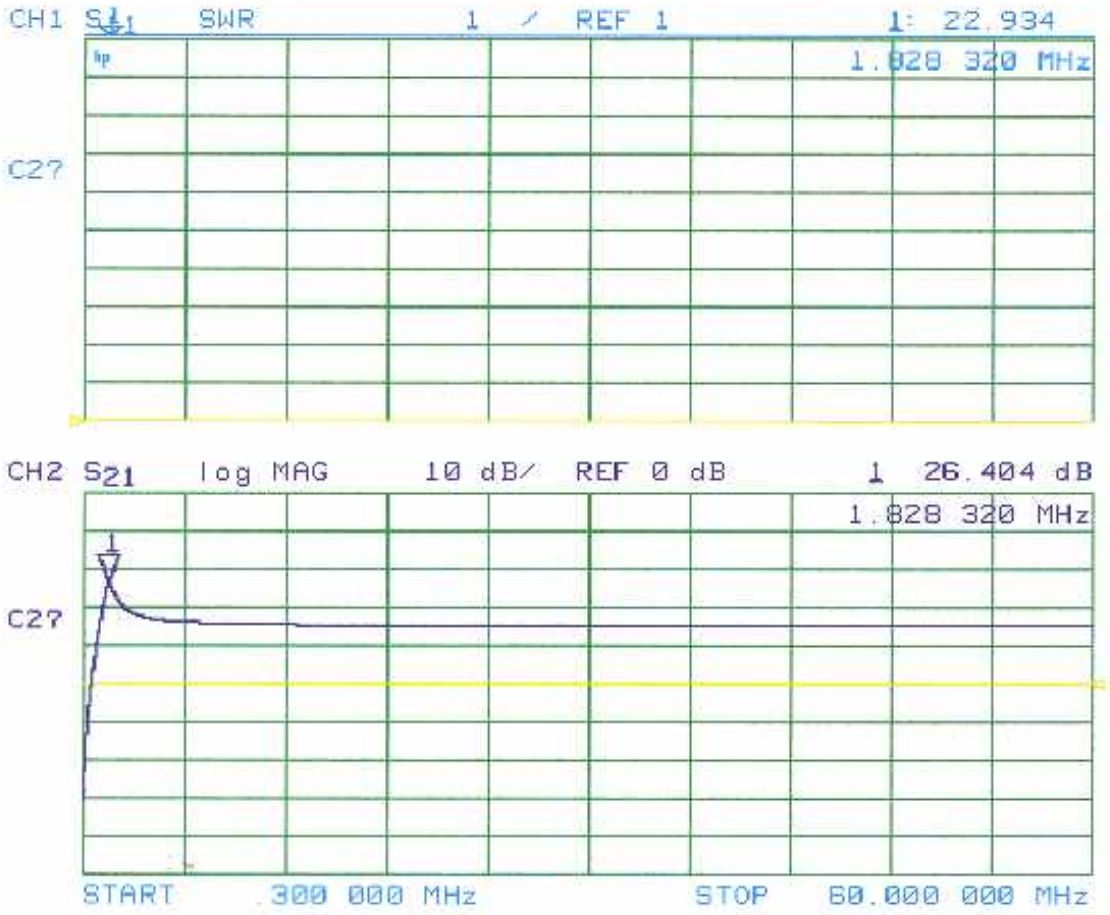


DXE Smith Chart (top) Gain (bot)
marker at 1.210 MHz.

This is the gain and input impedance at 1.207440 MHz.

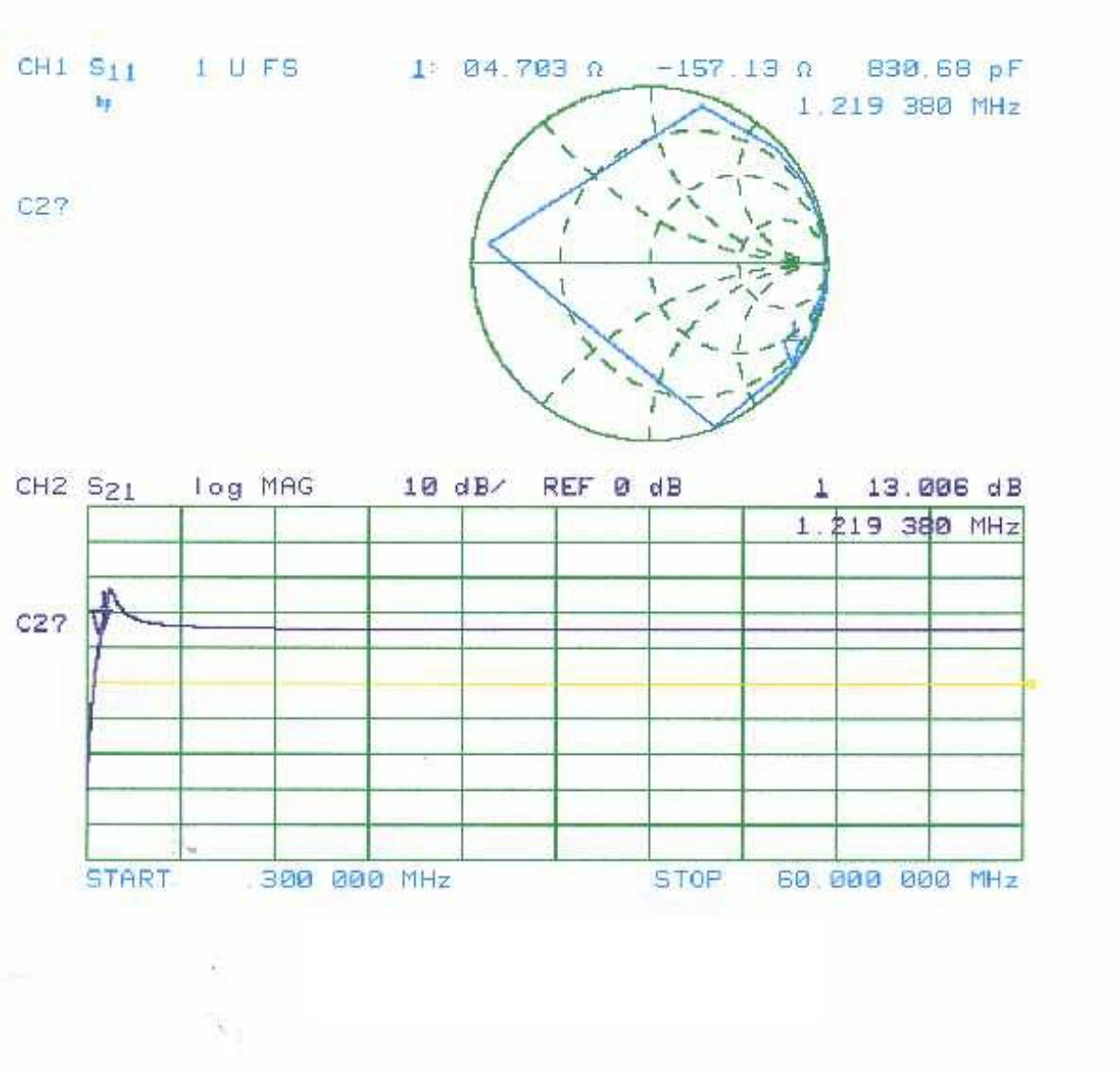
ARR GAsFET PreAmplifier

This amplifier is intended to be a 160-meter only amplifier. It has high gain despite of the very poor match on any frequency. It also has substantial gain on all frequencies swept, but a strong peak on 1.8 MHz.



ARR GAsFET SWR (top) Gain
(bot) marker at 1.828 MHz,

Input SWR is 23:1, gain is 26.4dB. Unfortunately noise figure is not that good for a GAsFET, probably because of the very large input



mismatch.

The above data is a Smith Chart display of the ARR GAsFET at 1.22 MHz. Gain is still 13dB, input impedance is 5 ohms and very reactive (this measurement is in the BCB). Despite being a "160-only" amplifier this amp has substantial gain (~15dB) at 60MHz.

Other Data

If you have a commercial amplifier or ***commonly available*** homebrew amplifier design you would like tested, please contact me. I'd be happy to measure your amplifier (time permitting) and post the results here.

See (soon to be added) summary page for text comparisons.

Amplifiers

[Home] [Amplifier Power Parts] [Bandswitch Failures] [Demonstration] [Pre-amplifiers]
[Tank Circuit] [SB-221] [Vacuum Tubes] [Vacuum Tube Amplifiers] [VHF Stability]
[FL2100 Problems] [HF Stability]

SB-221 Tank Voltages and the parasitic myth

Tuning or loading an amplifier

Preamplifiers

Receiving amplifier measurements

Vacuum Tubes

Vacuum Tube Amplifiers how the PA vacuum tube converts dc to RF

Safe working model for understanding how PA RF voltages are created and how amplifiers behave under various loading and drive conditions

Vacuum Tubes, arcing, vacuum tube failures, advantages and disadvantages of each type of power-grid vacuum tube, vacuum tube grid failures, vacuum tube anode failures, maximizing vacuum tube life, and quite a bit about how power grid vacuum tubes actually function

Components

Tank Circuits two models showing tank voltages, currents, loss, and phase shift

Bandswitch Failures discusses the many common reasons for amplifier tank system switch failures

T/R relays and relay timing, including amplifier relay sequencing that will work with vacuum relays

Inductors while this deals mainly with loading coils, it is also true for amplifiers

Stability

HF Stability HF instability or instability near the operating frequency of an amplifier is by far the most damaging for of oscillation. This problem plagues amplifiers like the Collins 30L1, Dentron Clipperton L, and Yaesu FL2100.

VHF Stability Parasitics discusses causes of instability. Has spice models, pictures of resistors internal construction. Your replacement carbon composition resistor might not be a carbon composition after all!

Arcing Causes

Antenna and antenna system failures that cause power amplifier arcs

Poor or improper relay sequencing causing power amplifier arcs. This is a common problem in TL-922 Kenwood amplifiers

Improper Drive and Amplifier Tuning how incorrect drive and loading adjustments cause power amplifier arcs, especially with exciter power overshoot

Amplifier Tube Arcs causes and cures including real reasons why grid chokes fail. This is a common problem in the SB-220 SB-221, TL-922 and other amplifiers.

Future Articles

Choke selection will describe the proper selection and construction of RF chokes

More pages soon.....

visits since January 2004

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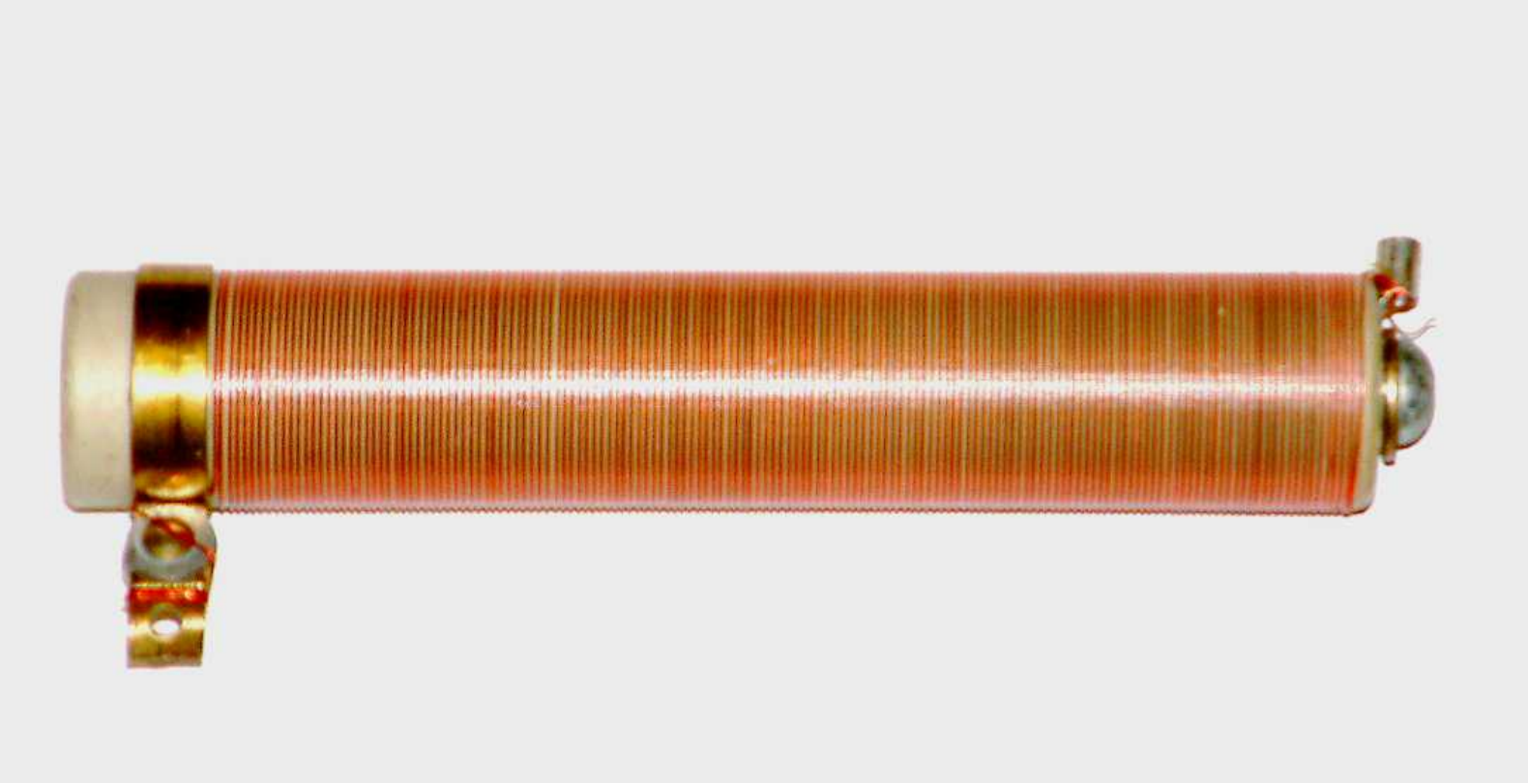
Amplifier Power Parts

[Home] [Up] [Vacuum Tubes For Sale]



RF Plate Choke

142uH @ 1.8 MHz. 3 amperes

Single layer space wound

Ceramic form

Hi-Q insulated

1” diameter by 5.5” long.

1/4" x 20 female thread both ends

Series resonances at 17, 25, and 32MHz. (Typically unusable at +-5% these frequencies)

Brand-new (similar Heathkit SB1000 and Ameritron)

$15.95 + 3.00 S&H in USA

Vacuum Tubes

I have a large stock of new vacuum tubes, including high power tubes.

Vacuum Tubes For Sale

[Home] [Up]

Transmitting Tubes

Various tube types from 100TH through 4CX15000 including hard to find 4CX1600B tubes.

List posted soon

Receiving Tubes

TUBE	QTY	BOX	TUBE	QTY	BOX	TUBE	QTY	BOX	TUBE	QTY	BOX	TUBES	QTY	BOX
35Z3	1	1	556	2	2	5687	1	3	1258	1	4	655	1	5
12SW7	1	1	1U5	1	2	6AF4A	11	3	10DX8	2	4	12A6	1	5
25L6	1	1	1X2C	1	2	6AG5	2	3	10GK6	2	4	12DQ6B	1	5
2A5	1	1	2AF4B	1	2	6AK5	1	3	10KR8	2	4	12SA7	1	5
50L6GT	1	1	2AV2	1	2	6AL5	1	3	11CY7	2	4	12SA7GT	2	5
5Y3WGTA	2	1	2CY5	3	2	6AQ5	3	3	11KV8	1	4	12SQ7	2	5
6JB6A	4	1	2FS5	1	2	6AS8	1	3	11LT8	1	4	1N5GT	1	5
6BG6A	1	1	2HA5	2	2	6AT6	1	3	12AB5	2	4	25L6GT	1	5
6BQ6GTB	2	1	3A2	1	2	6AT8A	2	3	12AD6	1	4	35Z5GT	2	5
6BX7	1	1	3AL5	4	2	6AW8	2	3	12AE7	2	4	50A5	1	5
6EW7	1	1	3AU6	3	2	6BC5	1	3	12AL5	2	4	50L6GT	2	5
6K6GT	1	1	3AV6	2	2	6BC7	1	3	12AT6	1	4	6BG6	1	5
6SG7Y	1	1	3BC5	2	2	6BC8	1	3	12AU6	4	4	6BY5GA	1	5
6SL7WGT	2	1	3BE6	1	2	6BE6	2	3	12AV6	31	4	6DQ5	1	5
6V6GTY	2	1	3BY6	1	2	6BF5	1	3	12AV7	3	4	6JE6A	1	5
6X5GT	1	1	3BZ6	1	2	6BH8	1	3	12AW6	1	4	6KN6	1	5
7C5	1	1	3CB6	2	2	6BJ8	2	3	12BA6	6	4	6KN6	1	5
6N060T	1	1	3CS6	2	2	6BK5	3	3	12BF6	2	4	6S57	1	5
1613 (JAN Number)	2	1	3CY5	1	2	6BR5	1	3	12BH7	1	4	6SA7	1	5

			3DT6A	2	2	6BS8	2	3	12DT5	2	4	6SA7	1	5
			3EH7	1	2	6BW8	1	3	12F8	1	4	6SC7	1	5
			3GK5	1	2	6BY8	3	3	12FX5	1	4	6SF5	2	5
			3HA5	1	2	6BZ7	2	3	12HL7	1	4	6SK7	1	5
			3V4	3	2	6C4	3	3	15HB6	1	4	6SK7	3	5
			4BQ7A	2	2	6CM7	1	3	17C9	4	4	6SN6	1	5
			4GK5	3	2	6CQ8	1	3	18FW6A	1	4	6SN7	1	5
			4HS8	1	2	6CS6	1	3	18FX6A	1	4	6SQ7	3	5
			4LJ8	1	2	6CS7	1	3	18FY6A	2	4	6V6GT	7	5
			5AM8	1	2	6CS7	1	3	19J6	4	4	6X5GT	1	5
			5AN8	2	2	6CW5	1	3	19T8	2	4	O24	2	5
			5AQ5	1	2	6CX8	1	3	20EZ7	1	4			
			5AT8	3	2	6DE6	1	3	25BK5	1	4			
			5BK7	1	2	6DE7	1	3	35B5	1	4			
			5BR8	3	2	6DK6	1	3	35W4	5	4			
			5BT8	1	2	6DT6A	2	3	50C5	1	4			
			5CG8	4	2	6EB8	1	3	50DC4	1	4			
			5CL8A	2	2	6EH5	1	3	50EH	5	4			
			5FV8	2	2	6EH7	1	3	684A	1	4			
			5KE8	1	2	6EH8	1	3	6JCA	4	4			
			5T8	1	2	6ES5	1	3	6JU8A	1	4			
			5U8	1	2	6ES8	1	3	6LC8	3	4			
						6EU8	3	3	6LE8	3	4			
						6FG7	2	3	6LN8	2	4			
						6FQ7	4	3	6LX8	3	4			
						6FV6	2	3	6T8	1	4			
						6GK5	1	3	6X4	1	4			
						6GK6	1	3	6X9	3	4			
						6GN8	1	3	6Y9	1	4			
						6GW8	1	3	8BQ5	2	4			

						6HZ6	1	3	8KA8	2	4
						6JC6A	2	3			

TUBES	QTY	BOX	TUBES	QTY	BOX	TUBES	QTY	BOX	TUBES	QTY	BOX
12AU7	1	6	32	1	7	80	1	8	0B2	1	9
12AU7 used	31	6	43	1	7	3223	1	8	10DE7	3	9
12BA6	8	6	45	1	7	5687	1	8	12AT7	5	9
12BF6	1	6	2526	1	7	5692	1	8	12SQ7	1	9
12X4	1	6	5744	2	7	5842	1	8	12V6	1	9
1K5	1	6	5933	1	7	6080	1	8	25C6	1	9
1S5	1	6	6087	1	7	6100	1	8	3Q4	1	9
1T4	1	6	117Z6	1	7	6136	1	8	50C5	1	9
1U4	1	6	12AU7	1	7	6201	2	8	5Z4	1	9
25BK5	1	6	12AV5GA	1	7	7061	1	8	6AK5	2	9
3Q4	1	6	12AX7	1	7	7061	1	8	6AK6	1	9
6AF4	1	6	12SH7	2	7	7586	2	8	6AQ5	3	9
6AG5	1	6	12SH7	1	7	7868	2	8	6AS6	1	9
6AH6	1	6	12SH7	1	7	10-4C BALLAST	1	8	6AU6	4	9
6AK6	1	6	14A7	1	7	12A56	1	8	6AV5	1	9
6AL5	2	6	14AF7	1	7	12AU7	8	8	6BA6	3	9
6AM8A	1	6	17D4	1	7	12AV6	1	8	6BE6	3	9
6AN5WA	1	6	25CD6GB	1	7	12B4A	2	8	6BH6	1	9
6AU6	2	6	3525GT	1	7	12BA6	1	8	6BN6	1	9
6AV6	5	6	50L6GT	1	7	12BE6	2	8	6BY5	1	9
6BF6	1	6	5651A	1	7	12BH7	1	8	6BZ6	1	9

6BN6	1	6	5V4	1	7	12DQ6	2	8	6CG8	5	9
6CG7	1	6	6080WA	1	7	12GN7	1	8	6CM7	5	9
6CM6	5	6	65U7	1	7	12SH7	2	8	6CM8	5	9
6EW6	10	6	6AS7	1	7	12SR7	1	8	6EJ7	1	9
6FG7	1	6	6AU6	2	7	12V6	1	8	6EU7	7	9
6GW6	5	6	6AU8A	1	7	17AX4GTA	1	8	6EU7	5	9
6J6	1	6	6BG6GA	1	7	17EW8	1	8	6J6A	2	9
6SU8	5	6	6BH6	1	7	1V2	1	8	6K6	1	9
			6BJ6	1	7	25BQ6	1	8	6LW6	1	9
			6CA7	4	7	2D21	1	8	6Q7	1	9
			6CB5A	2	7	35L6	1	8	6S57	1	9
			6CD6	1	7	35W4	1	8	6SN7	1	9
			6SK7	1	7	35W4	1	8	6T8	1	9
			6V6	1	7	3CY5	1	8	6V6	2	9
			7F7	3	7	3DF3	1	8			
			DELAY RELAY 26 V 300 SEC		7	50C5	1	8			
			DELAY RELAY E-1215		7	5749W	1	8			
			DELAY RELAY X-35		7	5U4	1	8			
			117 V DELAY RELAY 300 SEC	1	1	5Y3	2	8			
						5Z4	1	8			
						6140/423A	2	8			
						6AG5	1	8			
						6AG5	1	8			
						6AL5	4	8			

				6AN8A	2	8			
				6AQ5	1	8			
				6AQ5	2	8			
				6AQ6	1	8			
				6AV6	1	8			
				6AW8	1	8			

TUBES	QTY	BOX
1AD2	3	10
1AY2A	3	10
2AH2	2	10
2AS2	4	10
3A3C	2	10
3BW2	1	10
3DB3	2	10
3DG4	1	10
5DJ4	2	10

Bandswitch Failures

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[Switch Installation]

Bandswitch Failures

Bandswitch failures are caused by:

1.) Improper installation or wiring of switch

a.) Improper Wiring Techniques

b.) Improper contact Alignment

c.) Contamination-of or damage-to switch

2.) Improper relay timing or defective antenna relay

a.) Transmit contact closes ***after*** input contact closes

b.) Transmit contact opens ***before*** input contact opens

3.) Failures in antenna systems

a.) Intermittent arc, open, or short in a tuner

b.) Intermittent arc, open, or short in an antenna

c.) Intermittent arc, open, or short in external lightning protection devices

d.) Intermittent arc, open, or short in a feedline or connector

e.) Sudden large change in either SWR or antenna impedance

4.) Improper loading of amplifier

a.) Loading control set too far meshed (loading too light)

b.) SWR or antenna impedance beyond matching range of amplifier

5.) Excessive drive or drive power transients

a.) Exciter has transients or leading edge overshoot

b.) Operation power peaks higher than tuning peaks

6.) Excessive current

a.) Improper loading (loading too light)

b.) Improper connections

c.) Improper switch ratings

Failures_in_antenna_systems

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Any of the problems below can result in destruction of a bandswitch or other amplifier components.

Tuners

Antenna tuners are a common problem. Many tuners are over-advertised, and will not handle the power claimed. Some models also have poor or unreliable connections in inductors. Many tuners have poorly designed power meters with peak meters that don't indicate true peak power. If you look at the antenna tuner review in the February 2003 issue of QST, you will see power rating problems exist with many tuners.

User adjustments also greatly control tuner power ratings. People adjusting "T" network tuners often forget to use the maximum possible capacitance required to match the load. Using less than maximum capacitance will reduce tuner power handling capability.

A very brief arc or open circuit in a tuner can cause a momentary very large change in antenna tuner input impedance. This large impedance change reflects back into the amplifier, and if the tank is momentarily unloaded a catastrophic bandswitch or capacitor arc will occur.

The worse possible change is one that results in a condition of high impedance at the amplifier tank input, in which cases tank voltages can reach many times the normal operating values.

Always be sure your tuner has reliable internal connections, and is tuned correctly (maximum capacitance that allows matching the load). Be sure the tuner can handle the peak transient power from your exciter-amplifier combination.

At one point, I owned an ICOM 775DSP radio. The peak transient power from that radio, even when set to 50 watts, was over 300 watts! While that pulse would not show on a normal power meter, it clearly showed on peak storage meters. My Bird digital storage, a special Coaxial Dynamics peak meter for medical applications, When I drove an 8877 amplifier with that radio, the few milliseconds of peak driving pulse would trip the very fast grid-current-fault protection system in my amplifier.

When I advanced the loading to prevent the grid trip from activating, the transient peak amplifier out was over 5kW. This would cause my antenna tuner, a large old KW Johnson Matchbox, to momentarily flash-over! When the tuner would arc, the amplifier tank would arc.

The entire system would have to be able to handle the exciter's 300-350 watt peak transient (amplified through the PA gain) in order to not have reliability problems!

Antenna and Feedline Failures

Antennas can have arcing problems from bad connections or other defects. If an antenna fails to accept power or changes impedance intermittently, even if for a few milliseconds, it can trigger an arc elsewhere in the system. That point may even be in an amplifier tank system.

Traps can arc, as can other components like gamma capacitors or matching networks. Be sure antenna wires are not able to brush up against other objects in the wind.

Be sure all connectors and hardware are tight, and connections are dry and well-insulated. Normal UHF connectors, even though rated less, handle much more power than type N connectors. UHF connectors are much less likely to develop intermittent connections, and actually have LESS loss below 50MHz. The only potential drawback for HF use is UHF connectors are less water resistant than properly installed N connectors, but that can be cured with proper waterproofing.

Antenna problems will show if you use a peak-storage meter, or a meter with a long hang-time. The only amateur radio products I'm aware of that have enough response time and hang time are the Ameritron meters used in the AWM-30 meter and ATR-30 and 20 tuners. Any momentary faults in my antennas immediately show on the reflected power of those meters, when they are in the peak mode.

Loading_amplifier

Improper Loading of Amplifier

As far as tank voltages are concerned, there really isn't any difference between excessive drive, load faults, or improper adjustment of loading. Improper tank adjustment and excessive drive are equally harmful to component life, and either can create splatter (and in extreme cases cause clicks) on adjacent frequencies.

Most Common Tuning Error

The most common tuning error is adjusting the amplifier at low power, rather than tuning at maximum obtainable drive.

The last few steps in tuning should be:

- Load the amplifier to maximum obtainable output at full exciter drive
- After that, advance the loading control slightly beyond (towards less capacitance).

ALWAYS load your amplifier for maximum obtainable power, and reduce drive to rated or desired power levels! This ensures minimum voltage and current in the tank, and maximum possible linearity for the best signal quality. High grid current is a good indicator of excessively light loading in grounded grid amplifiers.

Exciter Transients or Power Overshoot

Maximum available carrier drive might not result in sufficient drive when an exciter has transients or power overshoot. Transients or overshoot only appears when going from zero power towards full operating power levels, a condition met at the leading edge of speech or CW transmissions. Once the ALC comes up, transients and/or overshoot will generally disappear. With transients or overshoot, it is impossible to tune your amplifier properly with either a carrier, tuning-pulser, or speech waveform. We can not just tune for maximum output and expect the amplifier to be properly loaded when the exciter has transients!

Transients and overshoot, being of short duration and infrequently occurring, make it impossible to tune correctly at maximum drive.

Let's assume your exciter is rated to deliver 100 watts, but has momentary peaks or transients of 160 watts while the ALC or power control loop "takes hold". Power surges of 160 watts, too short to register on normal power meters, occur at the start of ever transmission. Of course, if you don't run the exciter wide open, and reduce power to 50 watts, this problem is rally aggravated!

In this case, the maximum power setting of the exciter should be used while tuning. If you suspect or know the exciter has transient problems, the loading control should be advanced a reasonable amount beyond (further open) the maximum output power setting. This will allow the amplifier tank system to handle transients without arcing or component failure.

Easy-Drive Linear Amplifiers

Some people and manufacturers tout low drive as an advantage, claiming it offers "cleaner signals". Nothing is further from the truth.

Exciters almost always provide the best IM performance when operated at an average peak power a reasonable amount below full output, rather than very low levels. At low levels, exciter performance is dominated by cross-over distortion caused by low-level operation of stages optimized for higher power levels. ALC also bias back early stages, often increasing distortion in those stages. At very high levels, gain compression or bias shift becomes an issue. Exciters typically do best when operated in the area of 60-80% of rated power.

Worse yet, low drive amplifiers are especially susceptible to damage from exciter overshoot or transient problems. Transients and overshoot remains almost the same level regardless of exciter power control settings. As exciter power levels are reduced, the percent of power overshoot becomes worse.

The most undesirable situation is one where exciter power greatly exceeds an amplifier's normal drive power limit. Not only does this reduce system IM performance, amplifier drive transients are aggravated. Amplifiers should be designed or selected to match the exciter's maximum power output, or an external attenuator used to bring the amplifier's drive requirement up to the exciter's full power level.

Demonstration

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A Practical Demonstration

This test circuit is easy to duplicate, and demonstrates the extreme voltages that appear in an amplifier when the load is improper for the amount of drive power applied.

Years ago, almost every amplifier had a output transformer. The transformer was similar to tanks circuits in RF PA's, in that it transformed or matched impedances. Even though Q was very low, the transformer could still store and release energy.

Experienced people knew if the volume was turned up too high in a transformer-coupled output stage, and if the load was absent, voltage across the transformer would soar to many times the normal operating voltage. No load or grossly mismatched loads often resulted in damaged transformers, blown output devices, or other output circuit component failures.

Failures induced by load or matching faults would occur in conservative amplifier designs, where components would last years in continuous ***proper*** operation. An audiophile would NEVER think of operating his expensive tube-type amplifier at anywhere near full volume into an open load, let alone a load where grossly improper impedances are selected in amplifier output-transformer taps.

Blame was never placed on stability. Everyone knew and understood conservatively designed well-constructed amplifiers with energy storage systems of any type, even very low Q systems, would still produce over-voltage failures when grossly overdriven or misterminated.

TV manufacturers used this well-known effect to advantage in horizontal output sections, where a flyback transformer with moderately low Q would produce many times the actual turns ratio in peak voltage, because of energy storage. Even modern switching supplies and our automobiles depend on energy storage to produce entirely new voltages, far above supply voltage, without requiring parasitics or high Q.

RF Systems

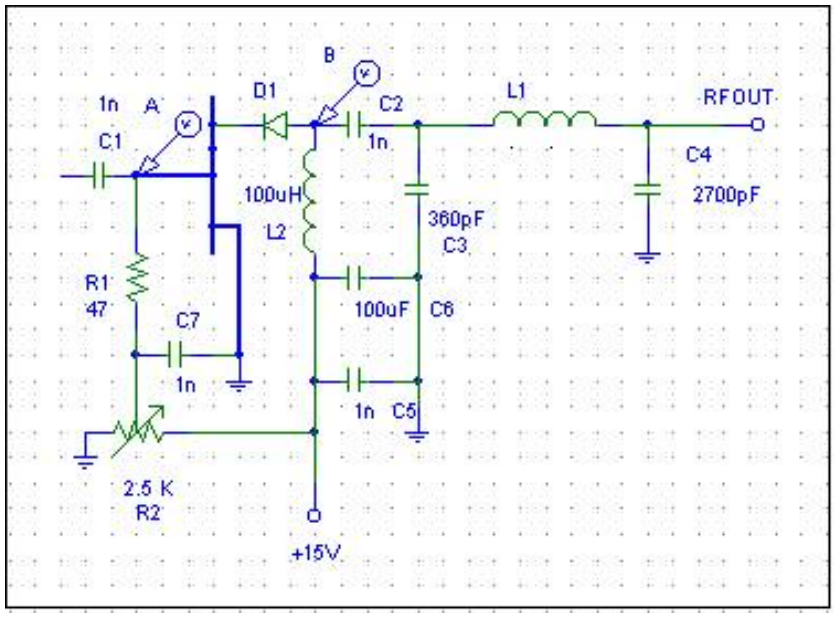
RF systems are certainly no more immune to mismatch than audio amplifiers, they very often are much worse. RF circuits are generally single-ended, and tube-type amplifiers have moderately high-Q (efficient) energy storage tank systems. Single-ended amplifiers with conduction angles under 360-degrees almost always contain intentionally designed “fly-back” systems, where tank circuit Q re-creates the missing portion of a sine wave from the half-cycle (or less) tug of the single-ended output device.

Somehow we have forgotten all this, and allowed ourselves to be misled into believing it takes a circuit or design flaw producing an oscillation to cause an arc or component failure.

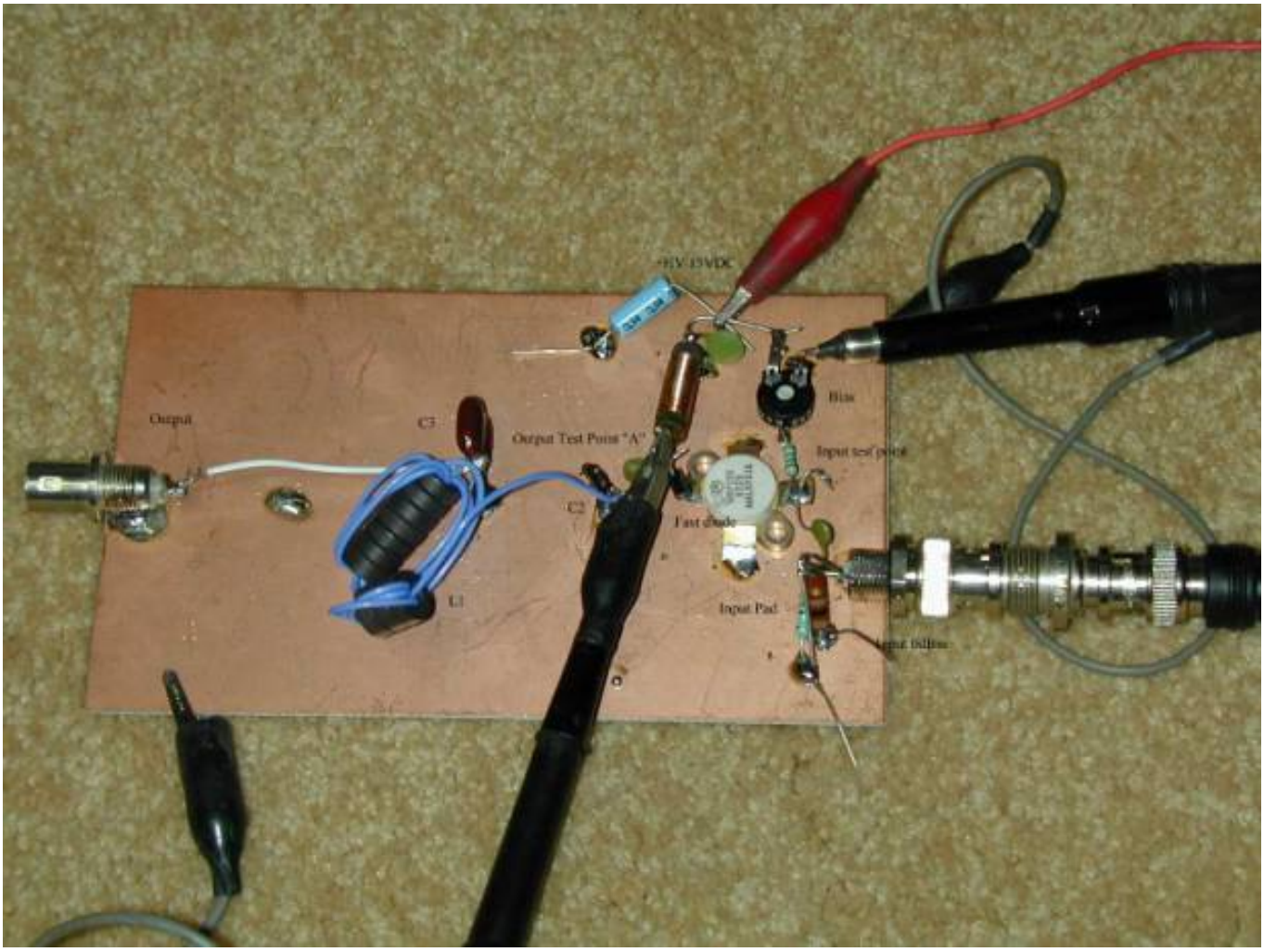
This article demonstrates how easy it is to produce very high voltages from normal perfectly stable PA's with normal tank systems.

Demonstration Circuit

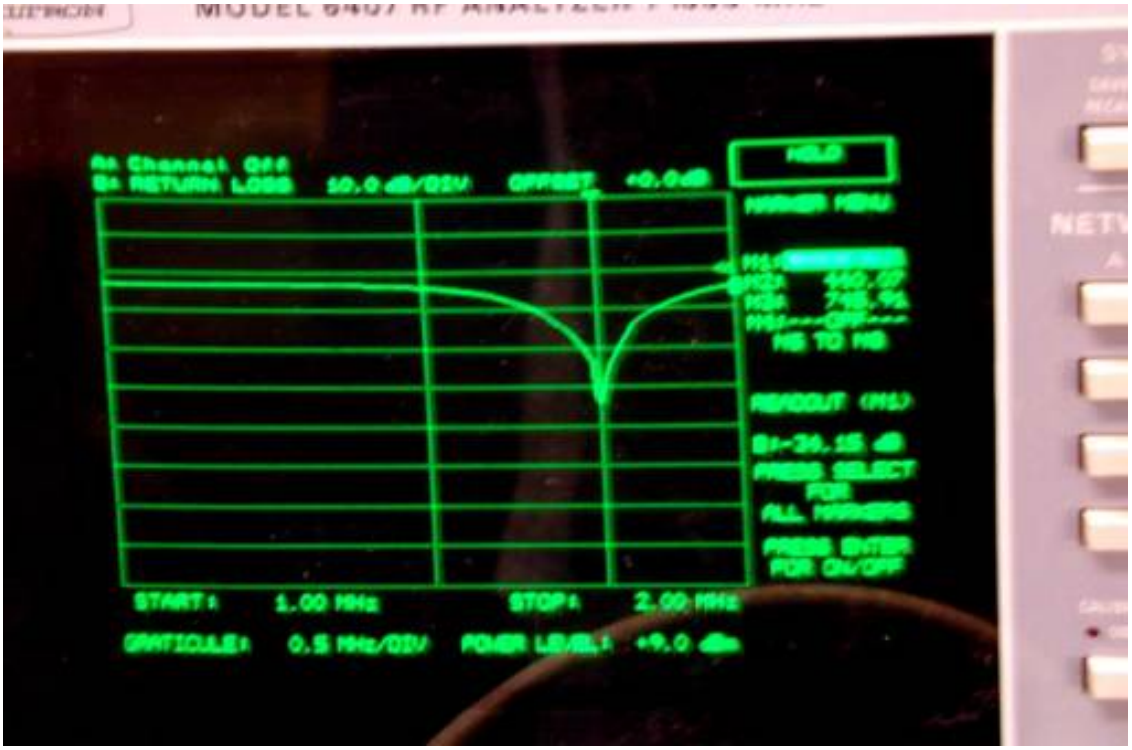
It isn’t safe to poke around in a high-power vacuum tube amplifier while looking at voltages, but a simple demonstration circuit can be constructed.



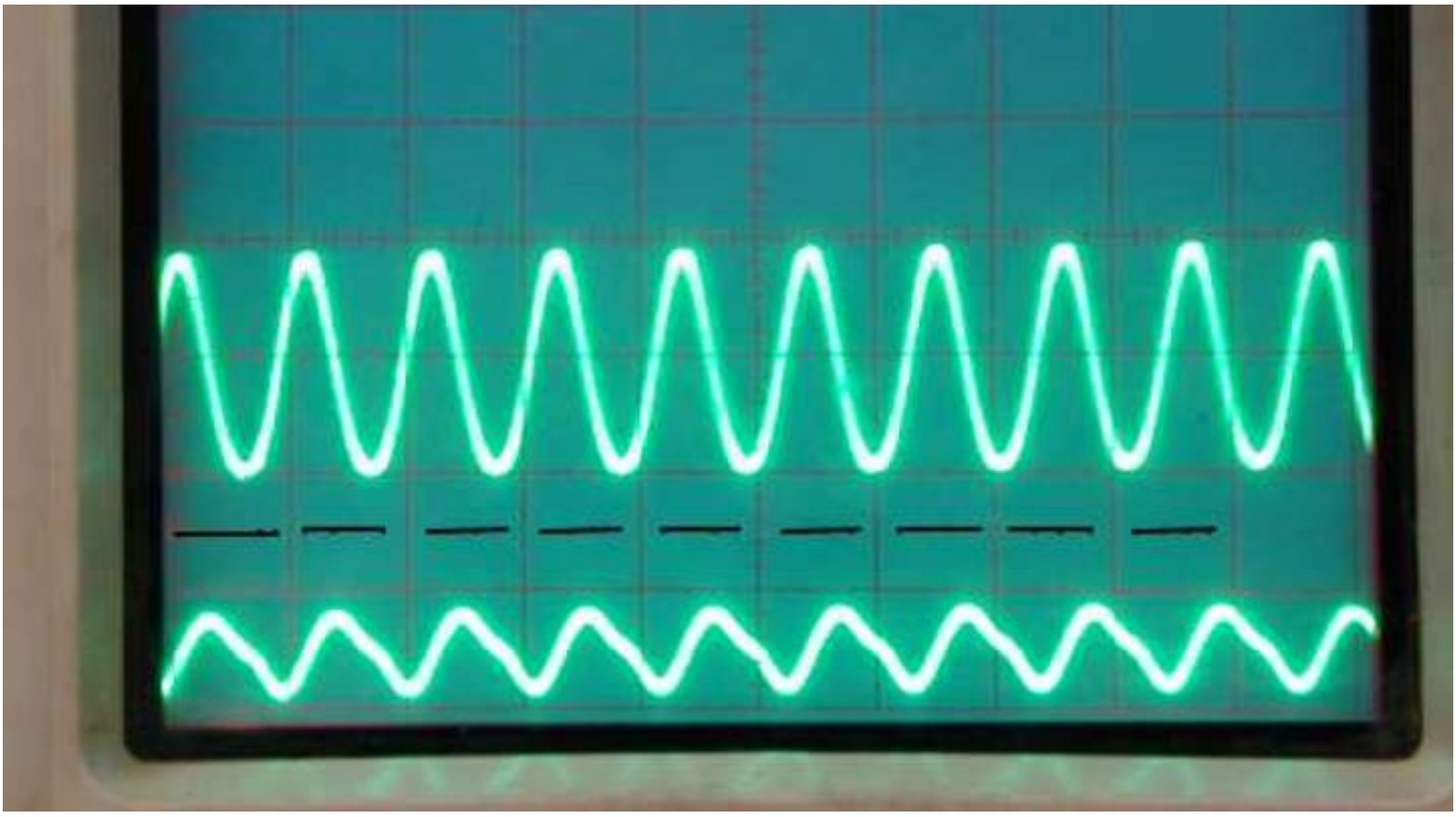
C1 is driven with a signal generator, L1 is selected to match the FET output to a 50-ohm load. The FET is operated at low current and has a series fast-switching diode, to simulate a one-way conducting vacuum tube.



This system was matched at 1.8MHz using the return-loss function of a network analyzer.



0dBm drive was applied with a signal generator, and the resulting waveforms appeared:

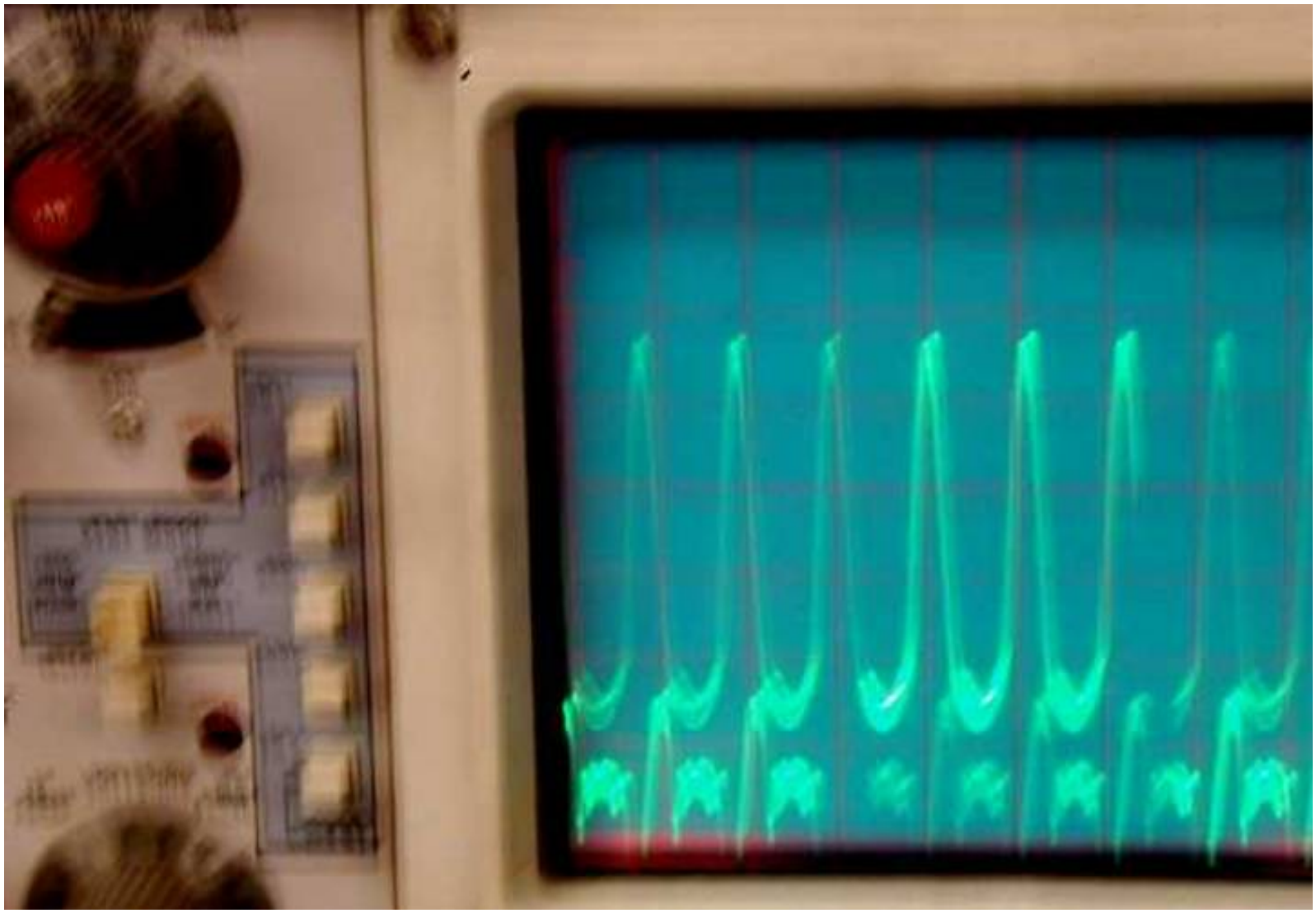


The upper trace is the output at D1 anode (point “B”). The black-dashed line was set at zero-volts. The scale is 10v/div.

The lower trace is drive voltage (point “A”). The scale is .1volt per division.

From this we see peak drain voltage swings approximately 20 volts, from around 5 volts to around 25 volts. This would be normal operation of a PA stage.

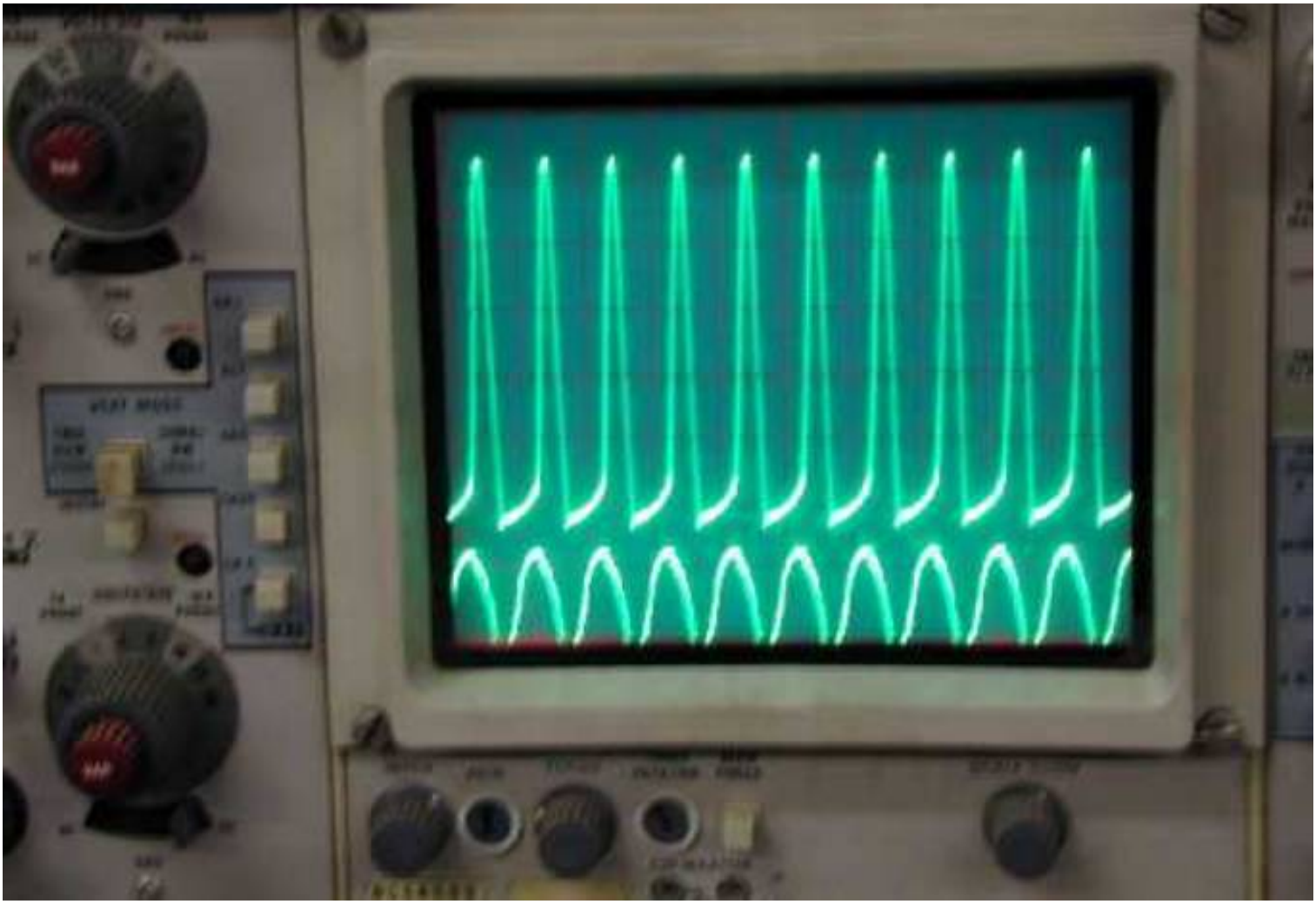
If we increase drive power and overdrive the PA, we get the following voltages:



We now see the peak drain voltage is almost 50-volts from a 15Vdc supply! In a 3000V PA stage, this would be 10,000V peak! It should make sense that amplifiers arc from grossly excessive drive power.

The next question would be what makes a normally driven amplifier arc. Often is when the load is inadvertently disconnected, either through poor relay timing, a bad cable or connection, or perhaps a failure in a component between the antenna and the amplifier.

Here is a scope picture with normal drive, but the load disconnected:



The drive has increased slightly because of feedback through the FET, now that the drain is swinging wildly almost 70 volts, all from a 15-volt supply. This would be the electrical equivalent of 14kV on the anode of a 3-500Z operating from 3000Vdc!

Conclusion

It’s easy to see why perfectly stable HF amplifiers, if overdriven or operated at moderate drive levels under conditions of a load fault, can be damaged by severe arcing.

Virtually all PA arcs (other than those caused by component failures or reduced voltage breakdown from dust or contaminates) occur when the load is interrupted or mismatched, and the PA no longer transfers energy to a proper load.

The vast majority of PA failures are caused by improper operation or defective components, not by "strange events" that are unpredictable and non-measurable.

Tank Circuit

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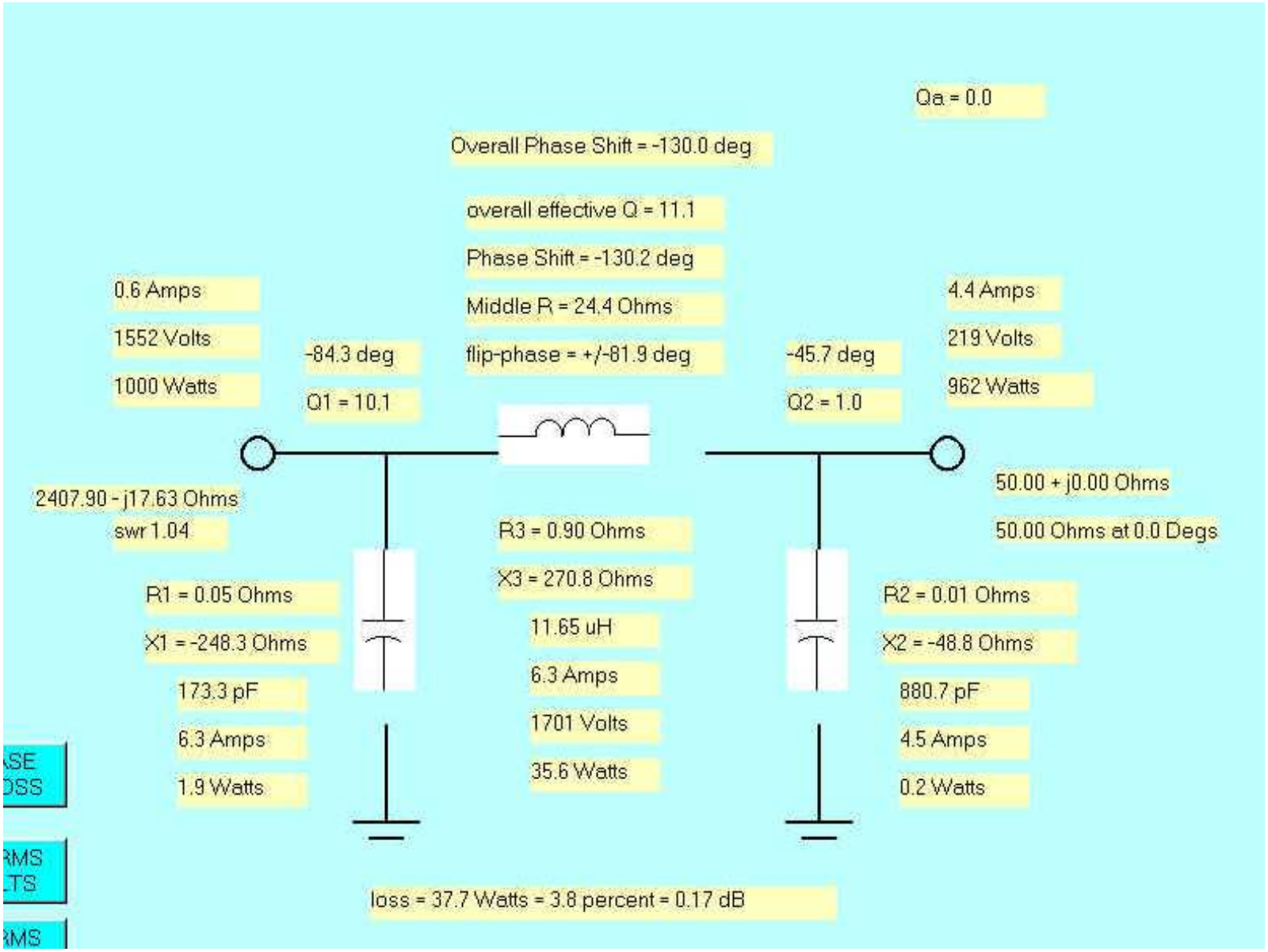
Amplifier tank circuit

The following is a model of an amplifier tank circuit with all parameters shown:



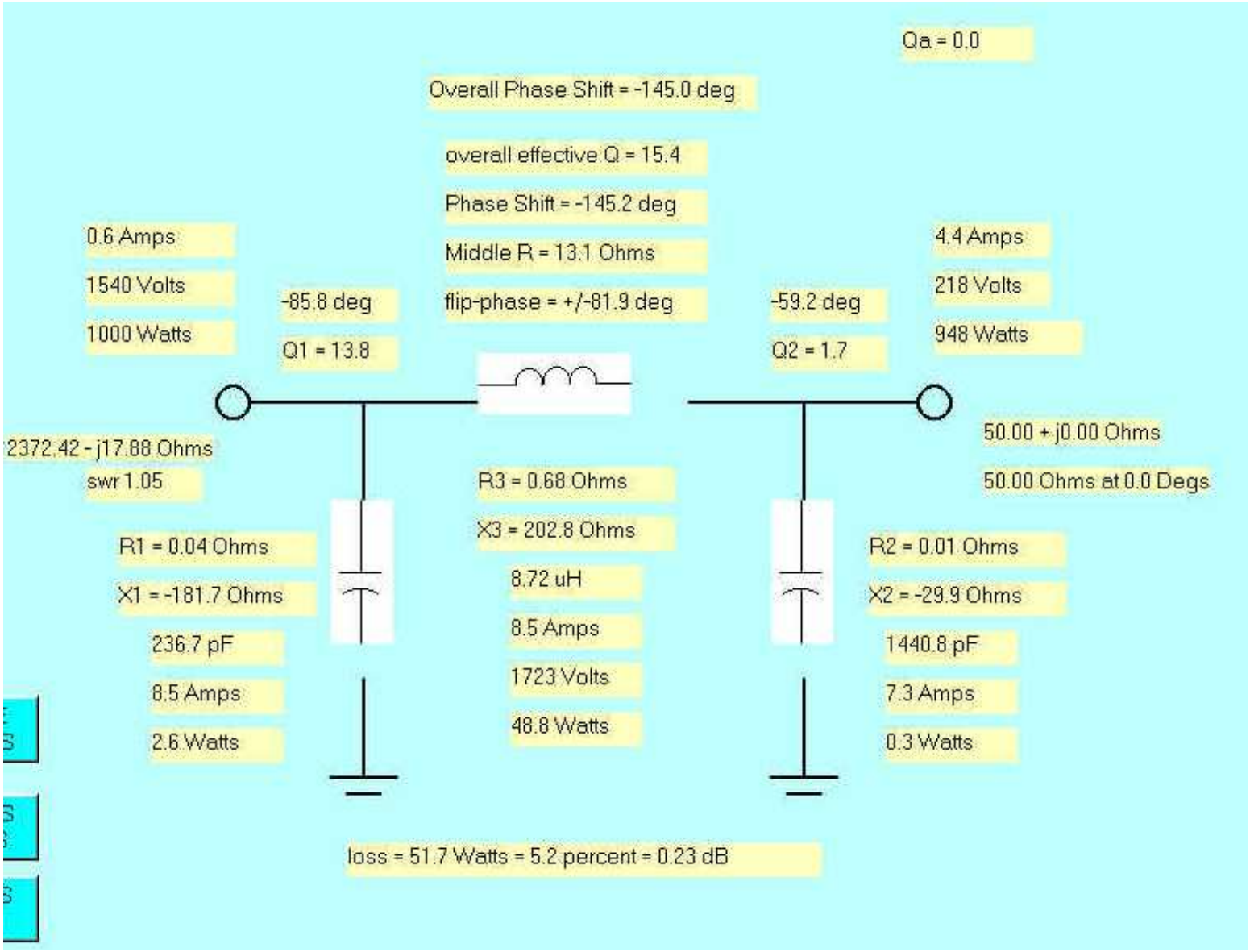
Notice the following:

Voltage across the tank coil exceeds input voltage

Current in capacitors exceeds input current

Phase shift across the inductor is -130 degrees

Increasing the loaded Q from ~11 to ~15 results in:



Notice the following:

1. Voltage across the inductor increases
2. Currents increase
3. Loss increases
4. Phase shift increases

Conclusions

(In all cases Q refers to loaded or operating Q)

1. Voltage across the inductor is always more than tank input voltage
2. Voltage increases with increasing Q
3. Phase shift increases with increasing Q
4. Currents increase with increasing Q

Voltages across a bandswitch contact can exceed tank input voltage, if the contact is near the input of the network. This, coupled with the high energy storage, can result in seemingly conservative components failing when the amplifier is lightly loaded for two reasons:

1. Tank circuit operating Q is higher when the tank is lightly loaded
2. The output device pumps the tank, and all of the energy may not be removed. This allows voltages at the input to greatly increase. See practical demonstration

Relay Timing

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Related page at **Relay Timing**

The text below covers electronic circuit to sequence and speed separate relays (like dual vacuum relays) and describes how to sequence mechanically ganged contacts like those in open frame relays.

Relay Timing and Sequencing

All amplifiers must be operated with a load on the output any time drive is applied to the input. Without this, the amplifier can be damaged by arcs or excessive internal voltages or currents. After all, all that power in the PA has to go somewhere! If there is no antenna connected, the energy builds up in the tank circuit or other energy storage system in the amplifier until something absorbs the excess energy that would have made it to the antenna.

In cases of unstable amplifiers, such as those using non-neutralized tubes with high internal feed-through capacitance (i.e. 572B's or 811A's), the output relay must be closed before any idling current is applied. (See stability)

The normal process of engagement sequencing is:

- The operator "keys" the radio (zero time)
- Amplifier output contact fully transfers (time "a" 1-10mS)
- The amplifier input contact fully or partially transfers (time "a" plus 1-3mS=time "b")
- Amplifier operating bias or cathode return path contact closes, applying normal operating bias (time "b" plus 0-2mS=time "c")
- Radio output delay, RF appears at the output of the amplifier (time "c" plus 0-10mS)

This entire sequence can take up to 15mS with large open-frame relays.

The release sequence must be:

- Transmitter ceases output of RF
- Bias drops
- Input relay drops
- Output relay drops

If this process is not followed the bandswitch, capacitors, or other tank components can be damaged from intermittent arcing!

Even though all radios should have a transmitter inhibit delay, many radio designs do not include such delays. The radio output delay primarily prevents shutting down of the radio during leading edges while relays transfer, and annoying broadband clicks or spits at the start of VOX closure or each transmitting cycle.

Some radios have incorporated delays, but incorrectly applied the delays. For example, some early Kenwood radios actually delayed the turn-on and turn-off of amplifier relays, forcing amplifiers to "hot-switch". Such radios are often disastrous to amplifier component life.

Increasing Relay Speed with proper sequencing/ Dual Relay System

Relays can be made faster by operating them from a higher-than-normal supply voltage, and using external current limiting to provide a constant current at the relay's rating. For a single relay, omit everything associated with RL2. R3 becomes a jumper.

A two-relay circuit would look like this:

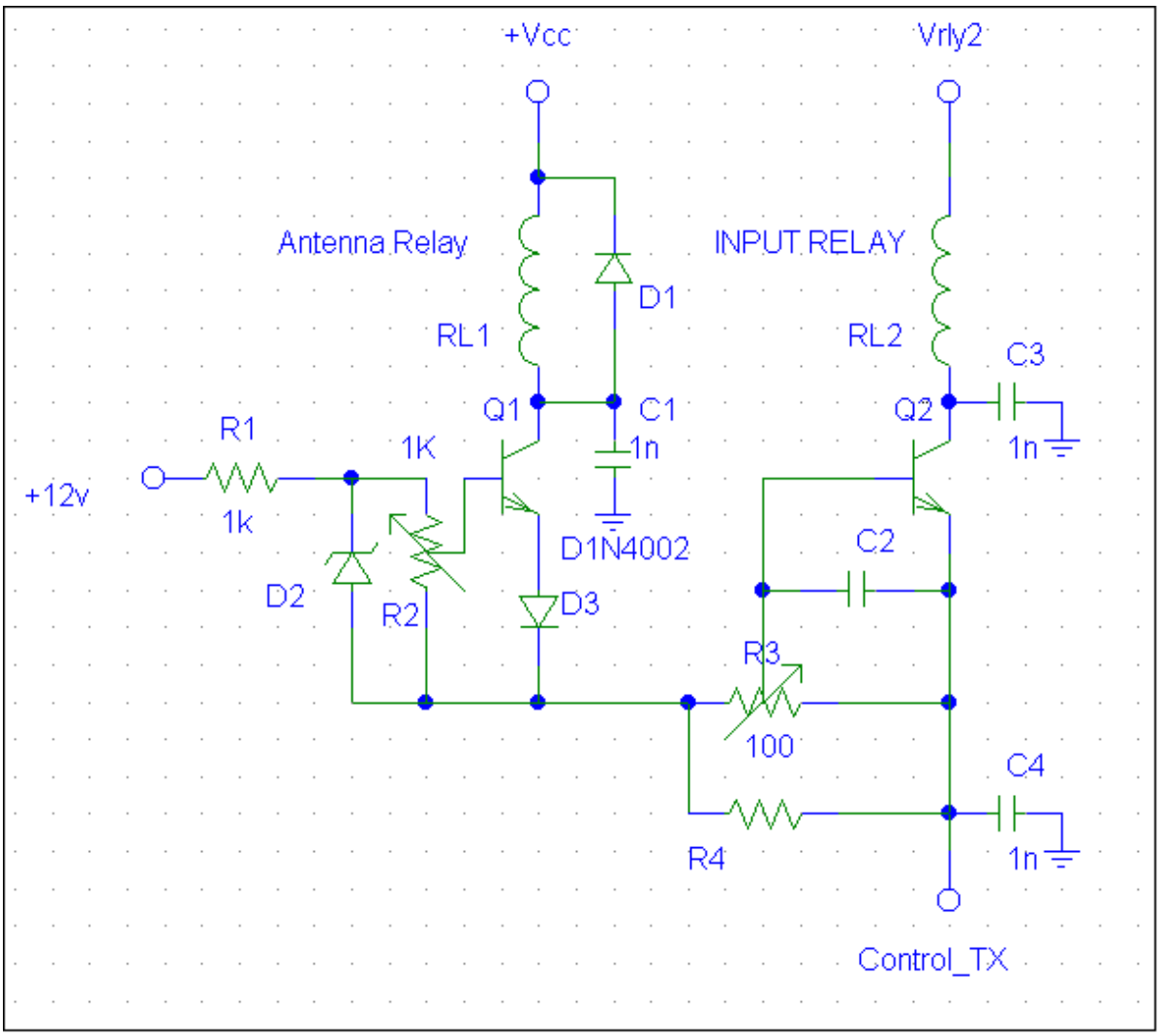


figure 1

R1 1k 1-watt

R2 1k 1/4-watt

R3 100-ohm 1/4-watt

R4 found by the following, where Irl1 is the antenna relay current:

$$R := \frac{.7}{\left(\frac{Irl1}{4}\right)}$$

For example:

$Irl1 := .25$

$$R := \frac{.7}{\left(\frac{Irl1}{4}\right)}$$

$R = 11.2$

Power rating of R4 is:

$P := I^2 \cdot R$ where I is RL1 current and R is R4 resistance.

C2 input time delay cap, normally 10-50uFd 10v

D1 1N4001-1N4007

D2 3-4V small zener (.25w-1w)

Q1,Q2 NPN power transistor. Vceo rating must exceed Vcc supply voltage, plus safety margin. Imax must exceed relay current ten times or more. Watch dissipation! Use hfe above 40 for best current stability.

Power Source:

- Vcc supply must exceed RL1 relay voltage by 2-3 times.
- Vrly2 supply should approximately equal RL2 rated voltage. This voltage must be obtained from a reasonably stable supply voltage +-20% regulation no load to full load. Do not use a large dropping resistor, or you will unintentionally speed up the input relay RL2.

R2 sets current through RL1 (antenna relay). The adjustment range of R2 is from zero current to a maximum where only RL1 resistance limits relay current. R2 must be adjusted for rated current through RL1.

R3 should be set for proper delay of RL2 (input relay).

D1 adds delay to release of RL1 (antenna relay). If release of antenna is too slow, add small value of series resistance until RL1 opens just after RL2 opens.

The above circuit has the following features:

- Control V open circuit will never exceed +12V
- No back-pulse or transient is generated
- Delays are easily tailored to any relay
- Extremely high or dangerous voltages are not present.
- Speed is faster than systems using dropping resistors for speedup
- Relay speed is maximized without exceeding rated relay current

How It Works

Relay pull-in is always slowed by the inductance of the armature coils. Since RL1 is fed by a voltage higher than normal, current will reach pull-in strength much sooner than with normal supply voltages. Q1 is a constant current source, adjusted by R2. This circuit prevents RL1 from being overheated, or having excessive steady-state operating current. Q1 also limits the voltage appearing at Control_TX.

After RL1 (antenna relay) current reaches a value that allows contact transfer, Q2 will conduct. This applies current to RL2, the input relay. It pulls in last, time delay is set by R4, R3, and C2.

Upon release of Control_TX, RL2 releases with only internal relay delays. RL1 is held on by the current loop through D1 as the field tries to

collapse. This causes RL1 to have a longer release time than RL2, so the antenna remains connected longer.

No back-pulse voltages make it to the exciter, because of the circuit configuration.

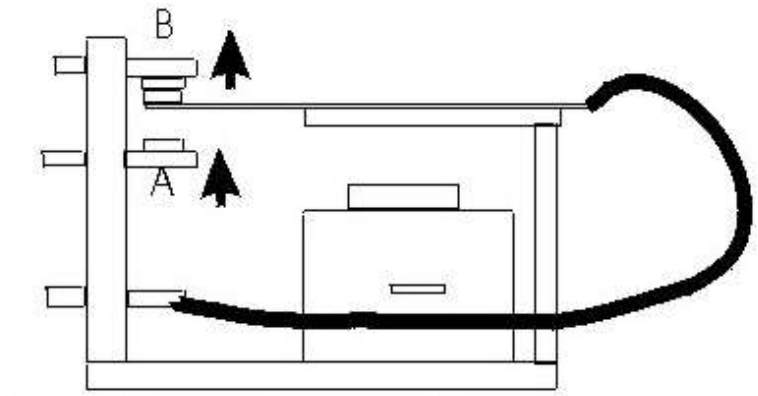
This is a simple, safe, adjustable relay system.

Mechanically Ganged Relay Contacts

Large multiple-contact open-frame relays using a common armature can be mechanically sequenced by slightly bending each contact carrier, the associated normally open contact, or both.

By carefully pushing on the armature center:

- The output contact should very obviously close first
- The input contact should close second
- The cathode or bias contact should close last, but nearly at the same time as the input contact



Relay Sequencing open frame relay

1.) Reference the pictorial above, use needle nose pliers to bend the unused normally OPEN contact at the amplifier tank circuit connection UPWARDS. Do this by positioning one jaw on the top of the bakelite frame and the other under the outer tip of "A" and squeezing until "A" is just slightly bent upwards towards the moving contact. ***DO THIS ONLY ON THE OUTPUT CONTACT!*** This will force the TANK CIRCUIT of the amplifier to connect to the antenna ***BEFORE*** the input circuit connects. **This change prevents arcs in tank components caused by improper relay timing.**

2.) Locate the unused normally closed contact for bias switching. Using the same technique above, bend "B" (the normally closed contact) upwards until B no longer contacts the moving contact. **This change puts 1/3 more contact pressure on the amplifier bypass contacts, and decreases receiver dropout when the amplifier is on bypass.**

When properly modified, you should see the ***output contact*** make slightly before any other contact on the relay when the armature is manually closed. You should see the UNUSED bias or cathode switching normally closed contact have a slight gap when the relay is NOT energized.

Separate Relays

Separate relays offer a unique problem. First, some relays can actually be dangerous to use.

NEVER USE REED RELAYS, UNLESS YOU ARE ABSOLUTELY SURE EACH RELAY HAS A SINGLE REED THAT MOVES BETWEEN THE NORMALLY OPEN AND CLOSED STATIONARY CONTACTS. MANY REED RELAYS HAVE A NO and NC REED IN A SINGLE COIL. THEY WILL RUIN EXPENSIVE RADIOS IF A REED WELDS SHUT!

NEVER USE COMBINATIONS OF SINGLE THROW RELAYS TO MAKE A DOUBLE THROW RELAY!!

DUAL REED RELAYS OR SEPARATE SINGLE-THROW RELAYS (like a SP NO and SP NC combined to make a SPDT section) CAN CAUSE SEVERE DAMAGE TO YOUR AMPLIFIER OR RADIO.

Relays must have an absolutely fail-proof positive transfer between open and closed poles. The normally closed poles should also be series-connected so the path is through both normally-closed contacts. This will prevent any possibility of accidental connection between amplifier output and input in the event of a relay or component failure.

Amplifier and Radio Relay Sequencing

[Home]

Related Page on Relay Timing

One problem on SSB and CW is relay time delay, especially when multiple devices are connected. Some radios have inherent RF output to external relay control timing issues. The result of improper timing is often a very wide leading edge "pop" or "click". The pop or click repeats at every relay closure. Slow VOX or semi-break in causes less frequent pops than QSK or fast VOX operation.

Required Sequence

The required sequence is:

1. Key closure or voice start
2. Unwanted delays in keying, VOX, and radio circuits
3. Amplifier relay or accessory (like front-end saver) signal activation
4. Desired RF hold off delay that is slightly greater than longest possible relay pull in delay for all switched devices
5. RF Appears
6. Keying or voice stops
7. RF stops (must include fall time of RF envelope)
8. Desired delay for complete signal fall before relay release

Time Delays

Most small to medium relays for amplifier switching are 5-15 milliseconds, with small enclosed and vacuum relays running around 5mS. Typical open-frame relays are in the 10-20mS range.

- **Relay closing time delay increases greatly as voltage across the coil is reduced. Decrease in speed can be as much as 300% as minimum operating voltage is approached.**
- **Relay opening time is delayed by excessively high relay holding current and back-pulse canceling diodes. The slowing can typically be as much as 200%. If you work QSK, use a design that eliminates back-pulse canceling diodes.**

PIN Diode switches can close and open in a few milliseconds or less.

Some very small reed relays rival PIN diodes in time. There are some low power reed relays that operate in less than 1mS!

Common Errors

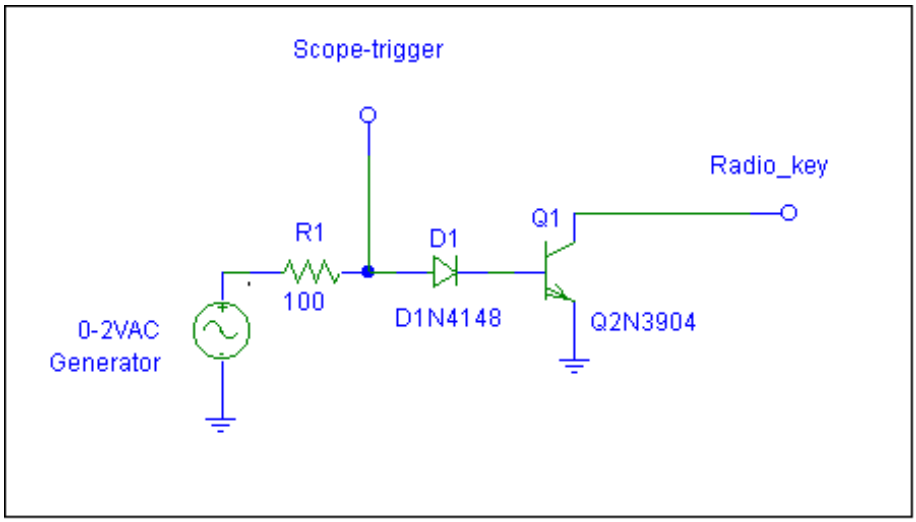
The most common errors are:

- **Lack of high operating voltage across the relay coil on pull-in. This is caused by excessive voltage drop in switching components or low relay supply voltage.**
- **Use of relays to buffer or isolate amplifiers**
- **Use of accessories that add time delay to relay lines**

Measuring Time Delays

Measurement of time delay requires a dual channel oscilloscope with separate trigger input, and a low frequency signal generator.

I normally use this keying test setup:



- The level of the generator controls weight or duty cycle (ratio of on-to-off time).
- The trigger is to the scope external trigger. The level threshold of trigger sets scope sweep starting time.
- The scope probes can be moved to any points to compare timing or sequence delays.

What to Watch

The radio must have enough RF output delay AFTER the radio low-amplifier control signal appears so the amplifier relay does not truncate

leading edge of RF

The radio or amplifier must hold the relay long enough so the falling edge is not truncated.

Problem Areas

- **Any device added between the control line output of the radio and the amplifier**
- **Any device that switches on the output of the radio**
- **Front end savers and amplifier relay buffers are a particular problem, often causing excessive control line delay times**

W8JI Control Line Interface

Some systems have multiple control lines from different devices that must be "keyed" in parallel. With direct parallel connections, one device can load the system and stop other devices from operating properly. This usually happens when a device is off but still connected. Simple isolation diodes work to prevent that, but the voltage and current at the radio is the highest voltage of any device and current is the sum of all the device currents. Even worse, the radio output control has to discharge the RF bypass capacitors of each connected device.

My 160 station consists of a solid-state homebrew amplifier with 16 MRF-150's, an AL1200, an 8877 amplifier and either an FT1000D or an Orion. I almost never listen through the FT1000 or Orion when working CW, so I also have at least two outboard receivers that must be muted.

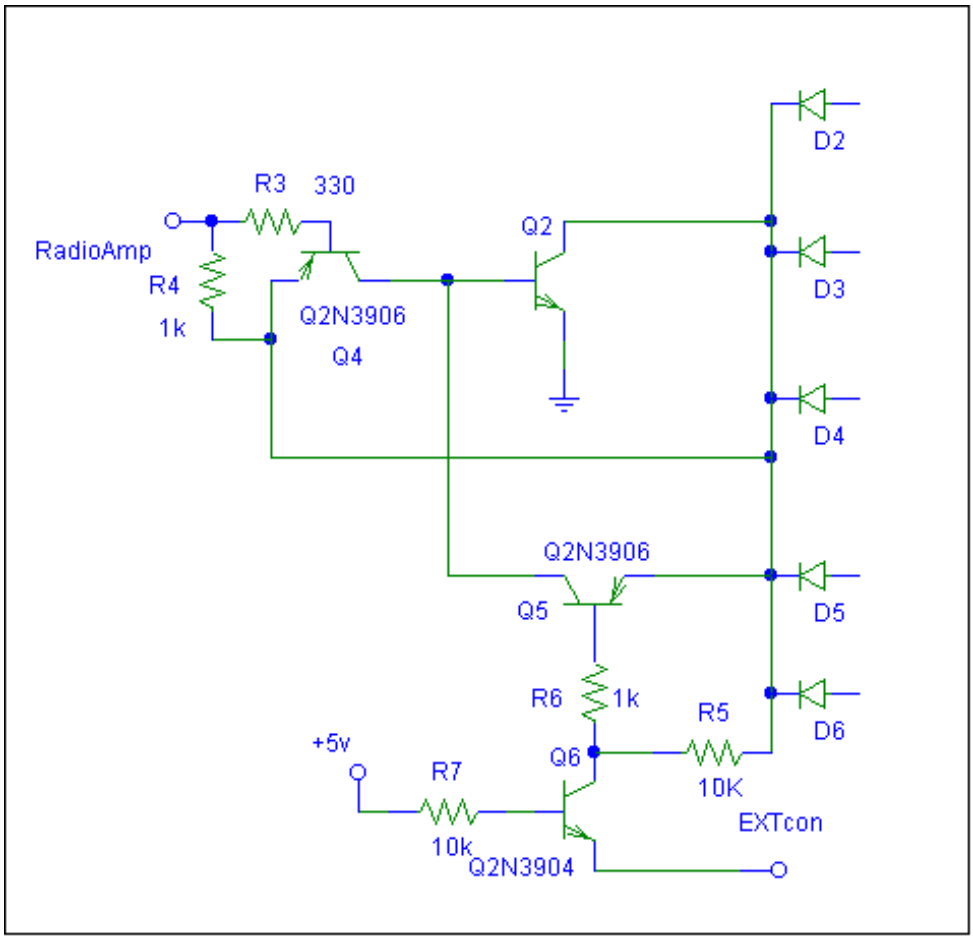
My homebrew 8877 amplifier uses a high-voltage relay system, the solid state amplifier has 5-volt TTL level switching, the TX antenna sampling line uses a 28-volt vacuum relay, and the receivers have 5 volt muting lines that mute when pulled to ground. The Ten-Tec Orion requires an isolation relay on the receive antenna port, and I occasionally switch other devices when transmitting.

All of this is driven by a microprocessor controlled antenna switching system that prevents changing antenna directions or antenna while transmitting.

Any attempt to directly parallel devices would be a disaster!

The circuit below interfaces my radio to multiple amplifiers, receivers, and antenna control systems. Multiple devices can be connected in parallel, such as 28 volt relay control lines and 5 volt control lines, as long as they are all positive voltage. This system pulls within 1 volt of ground.

It is necessary to be careful with some radios. There is no standard interface voltage. TenTec, for example, uses interface systems that have a threshold around 1 volt. TenTec's handshake lines require buffering with a small very fast reed relay, many transistors and IC's will not pull low enough to activate the TT handshake line. If you need relays of any type between the amplifier keying system and the radio, always be sure you test the entire system for proper relay sequencing.



RadioAmp goes to radio's amplifier keying control output.

EXTcon goes to computer or other keying device that only handles low-voltage low-current.

D2-D6 go to devices that require control.

+5V sets the threshold of EXTcon trigger. EXTcon open circuit voltage is also limited to less than 5 volts.

The radio is exposed to the highest voltage from the keyed loads, but the current is limited to roughly V/500 where V is highest voltage device. A 28 volt amplifier would require the radio to key about 14mA peak current during relay transition, although average current would be lower. The exact current depends on beta of Q2 and Q4, and if EXTcon is used. A device like Q6 can be added to the radio line to limit radio voltage to 5 volts or less regardless of amplifier relay voltages.

HF Stability

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HF instability, or instability at the operating frequency of the amplifier, is the most damaging type of problem. This is because the primary energy storage system, the tank circuit, is resonant with high Q at or near the operating frequency.

If you look at network analyzer swept measurements of a tank circuit, you'll see the maximum voltage distributed through the tank occurs at or near the operating frequency.

Some amplifiers have designed-in HF stability problems. The Yaesu FL2100 is one example of an amplifier with poor RF design. Un-neutralized amplifiers using tubes with high feedthrough capacitance are another source of problems. Any amplifier with high feedthrough (or feedback) capacitance is a candidate for HF instability if the load is removed while the amplifier tube or tubes are drawing quiescent current. Dentron amplifiers with four 811A and four 572B tubes, as well as the Collins 30L1 with four 811A's, are examples of relatively unstable amplifiers (along with the FL2100).

The cure is to remove intentional undesired feedback or to neutralize the PA!

SB-221

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SB221 Tank Circuit Voltage Distribution Vs. Frequency

Arcs in the bandswitch, tuning capacitor, and other tank components are often blamed on parasitics. No one ever can catch the amplifier in such a state of oscillation, so we are just left with conjecture and speculation that the arc was caused by a VHF parasitic.

It should be easy to prove if a VHF oscillation, or oscillation or drive on any frequency, could cause an arc. All we need to do is excite the anode of the PA tube with RF, and measure the voltage distribution in the tank with a very high impedance detector probe. This would then tell us at what frequencies the highest voltages occur.

While every amplifier manufactured is claimed to suffer from parasitics, I have limited time to measure and document the results. It would be impossible to test every PA, so I picked a PA that is commonly rumored to have parasitics.

The purpose of the series of measurements below is to see if it is possible for resonances in the tank to INCREASE anode voltage to the point where an arc will occur, or if it just an imagined problem or a problem "invented" to sell a specific cure.

Test Method

It is not practical to drive an HF PA from frequencies between 1 and 300MHz, so we have to excite the tank another way. The best method is to use a constant voltage source injected at a tube anode, and measure the voltage throughout the tank without disturbing anything. This can only be accomplished with a high input impedance detector probe, and a voltage-leveled source. Fortunately I have both available.

In order to be accurate, the tank must be adjusted normally. I did this by tuning the PA with a 100 watt exciter and a 50 ohm load. After tuning, the amplifier is disconnected from power mains and from the exciter.

The tube is purported to be the source of unwanted VHF oscillation energy. By injecting the anode with a known voltage, and sweeping the tank slowly over a wide frequency range, we can determine voltage distribution at any point in the tank system. By looking at the relationship to anode voltage, we can determine if it is even possible to arc a tank at frequencies far removed from the operating frequency.

Of course this would not prove an arc would occur, it would simply prove if it is possible for an arc to occur and at what frequency. The higher the voltage at a given frequency, the more likely an arc would occur. If the voltage is much lower, and arc at that frequency would be impossible.

The Measurements

The anode is excited with a swept power-leveled source feeding a bridging pad and the anode. The sweep is from 1 MHz to 300 MHz. The small coaxial cable and multicolored wire are from the high impedance J-FET source follower probe used to feed the second RF port of an Agilent network analyzer. In this condition the network analyzer acts like a sweep generator with a voltage detector.

- HP Network Analyzer source connected to tune anode at tube. Constant power source leveled at 10dBm at injection point load.
- 100k-ohm 3pF active voltage detector probe
- Transmission loss calibrated at zero dBm for mid-scale into the 100k probe as detector, across 2700-ohm resistor.
- All displays 10dB/div voltage, sweep is 1MHz to 300MHz.
- 1600 data points, 5 second sweep time.

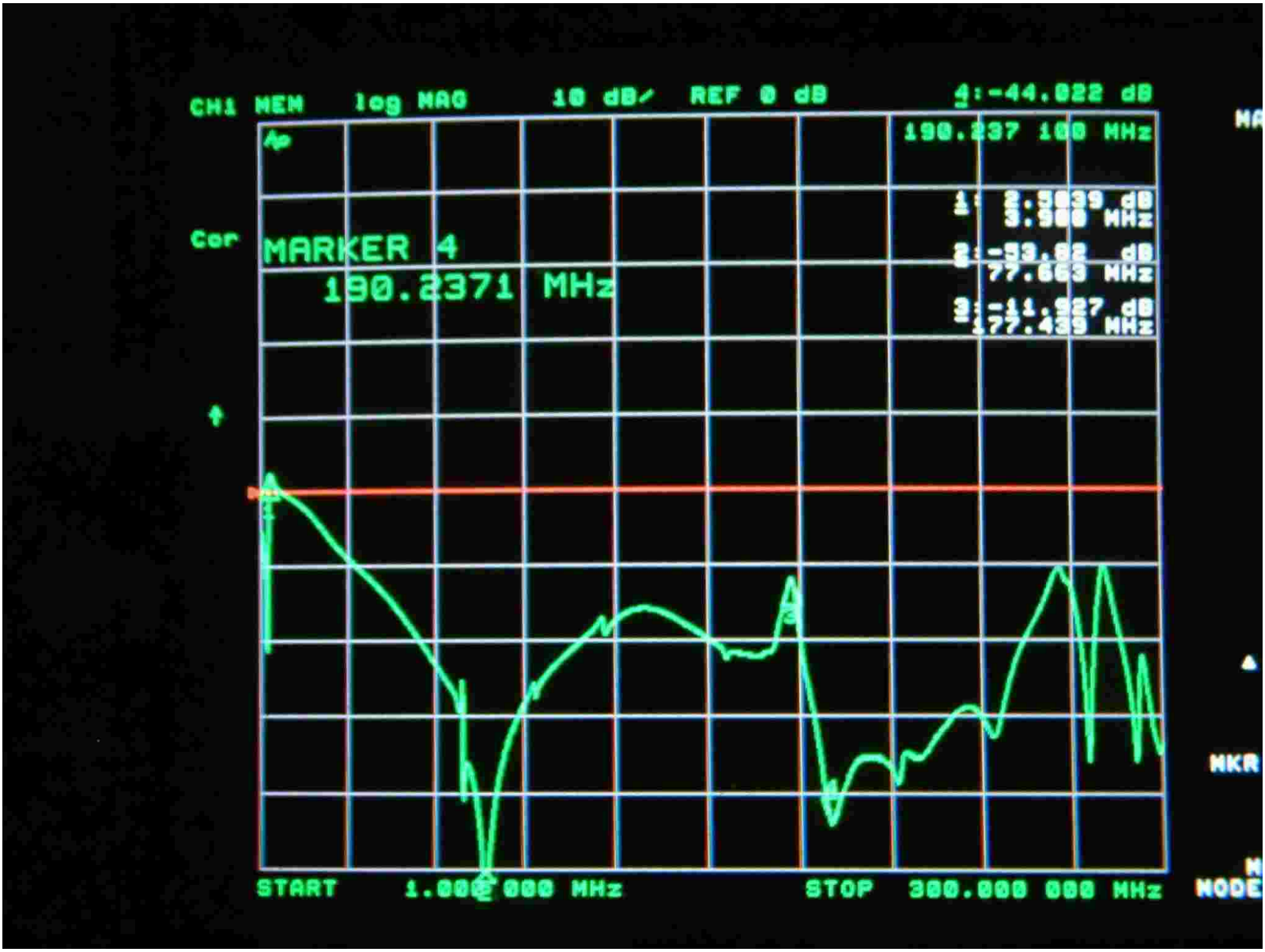

Figure 1. Tank adjusted for normal operation at 3.8MHz. The above sweep is voltage across the tuning capacitor, which is +2.5dB at plate tune capacitor. +2.5dB becomes reference for comparing HF voltages to parasitic voltages. With the anode at 3000 volts peak swing at 1300 watts on 80 meters, the reference level would be 2.5dB. A 1300-watt parasitic would develop the same starting voltage swing and it would be attenuated through the tank by the same amount as the displays indicate.

In this case at the tuning capacitor, a 180MHz parasitic would be 14.4dB down in level (marker 3) compared to the HF voltage (marker 1). This is 19% of the HF voltage level, or 570 volts peak with 3000 volts of anode voltage swing. A VHF arc at 180MHz would be impossible. As a matter of fact, there isn't any frequency very far from the operating frequency where an arc could occur!

A grid dip meter would show a strong dip at 180MHz, because it would be loaded by the tank. This is why a grid dip meter can not locate a "problem frequency", it might actually be locating a good frequency!

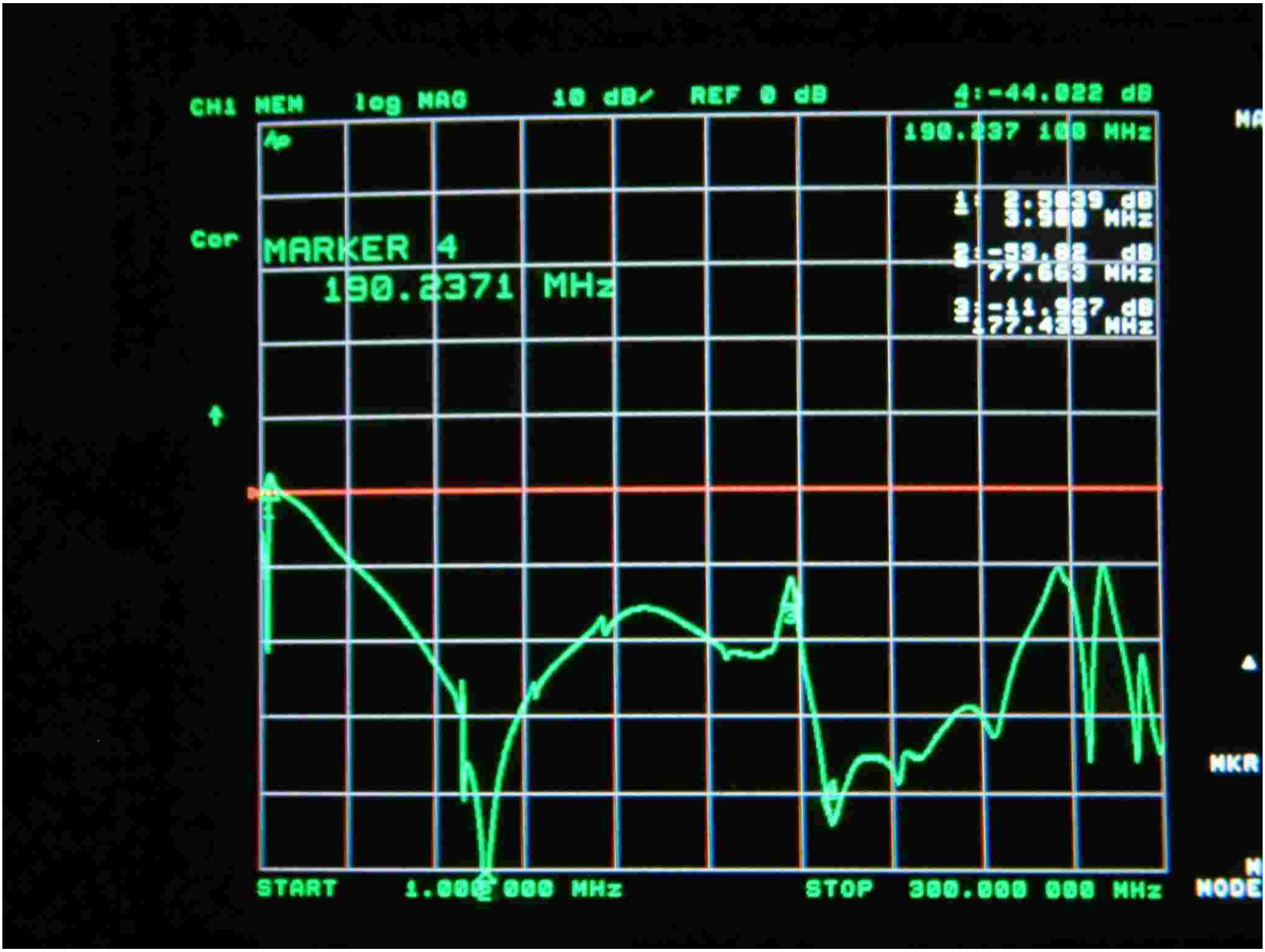

Figure 2. 80m position. The above graph indicates voltage at the 15 meter tap to closest arc point on switch. Highest voltage at frequency marker 3 is -12dB at 177 MHz. This is -14.5dB anode voltage, or 19% of anode voltage! 19% of 3000 would be only 570 volts, making it impossible to arc the 15M tap at VHF frequencies. After all, that tap survives constant exposure to over 3000v at the fundamental frequency when operating 80 meters!

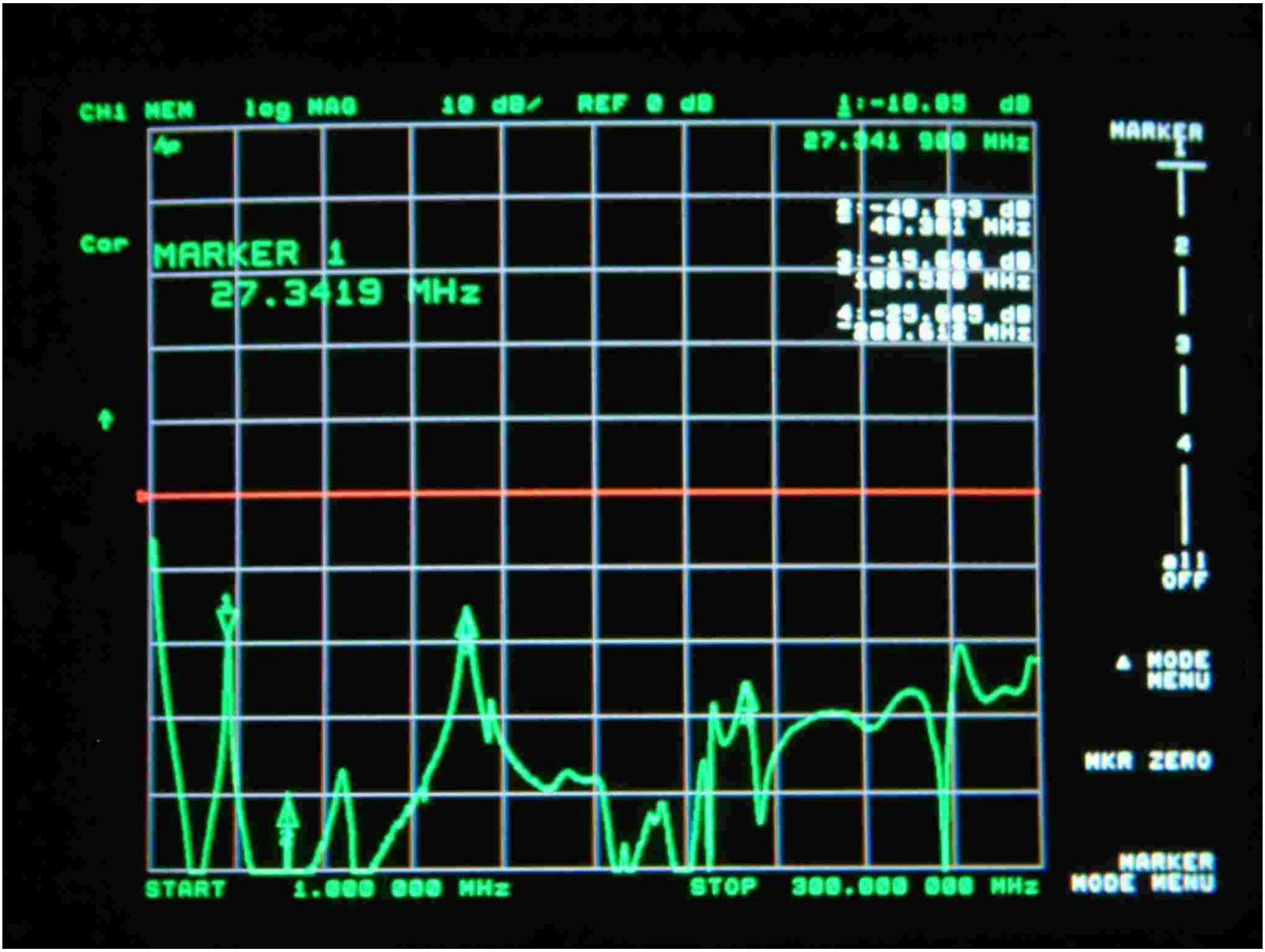

Figure 3. 80m position. Voltage across loading capacitor. All spurious response voltages are at least 20dB down from the anode voltage. No VHF arcs possible here either.

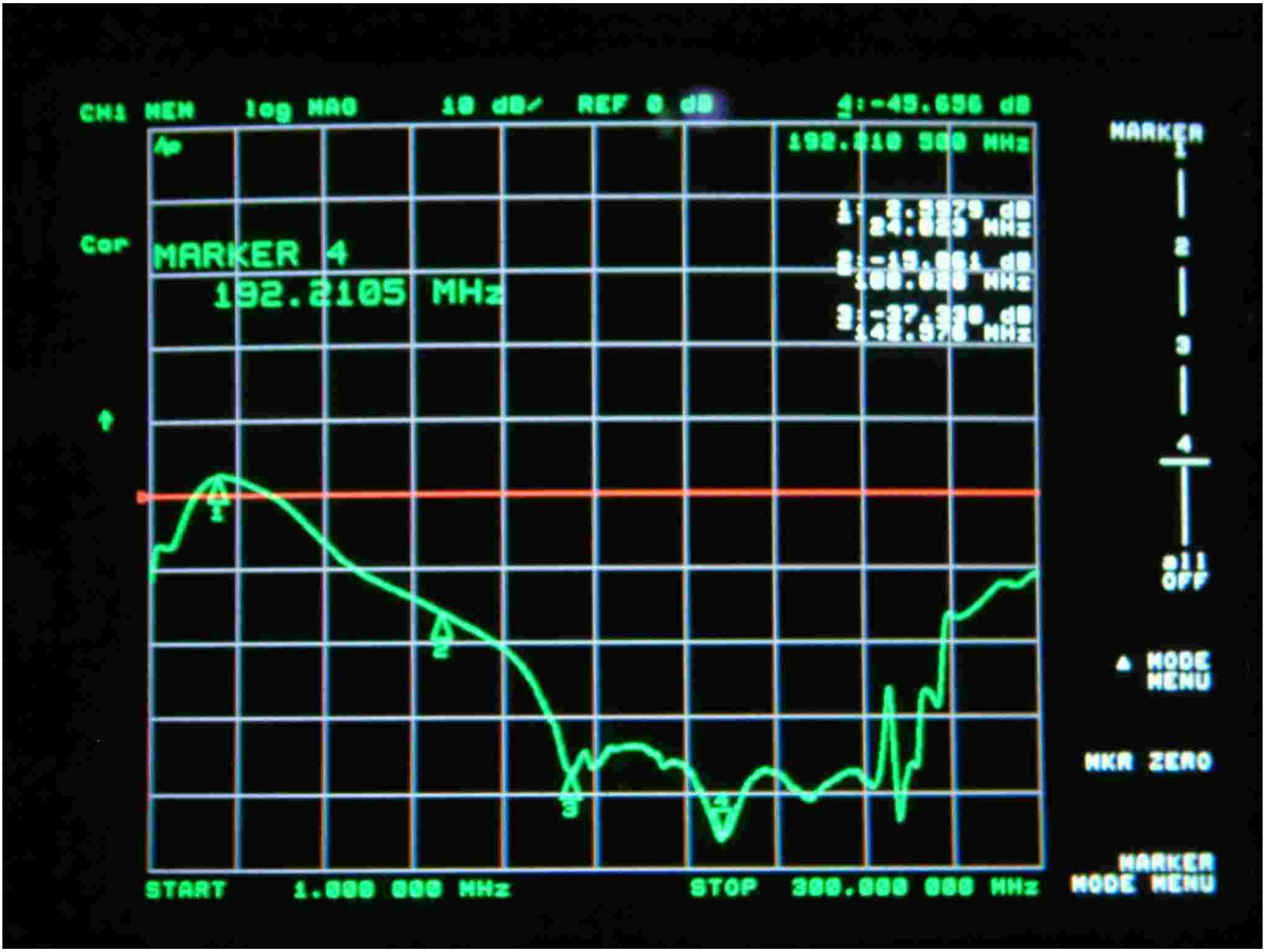

Figure 4. Tank normally loaded on 15-meters. Voltage across tuning capacitor. Anode injection voltage reference at +2.6 dB.

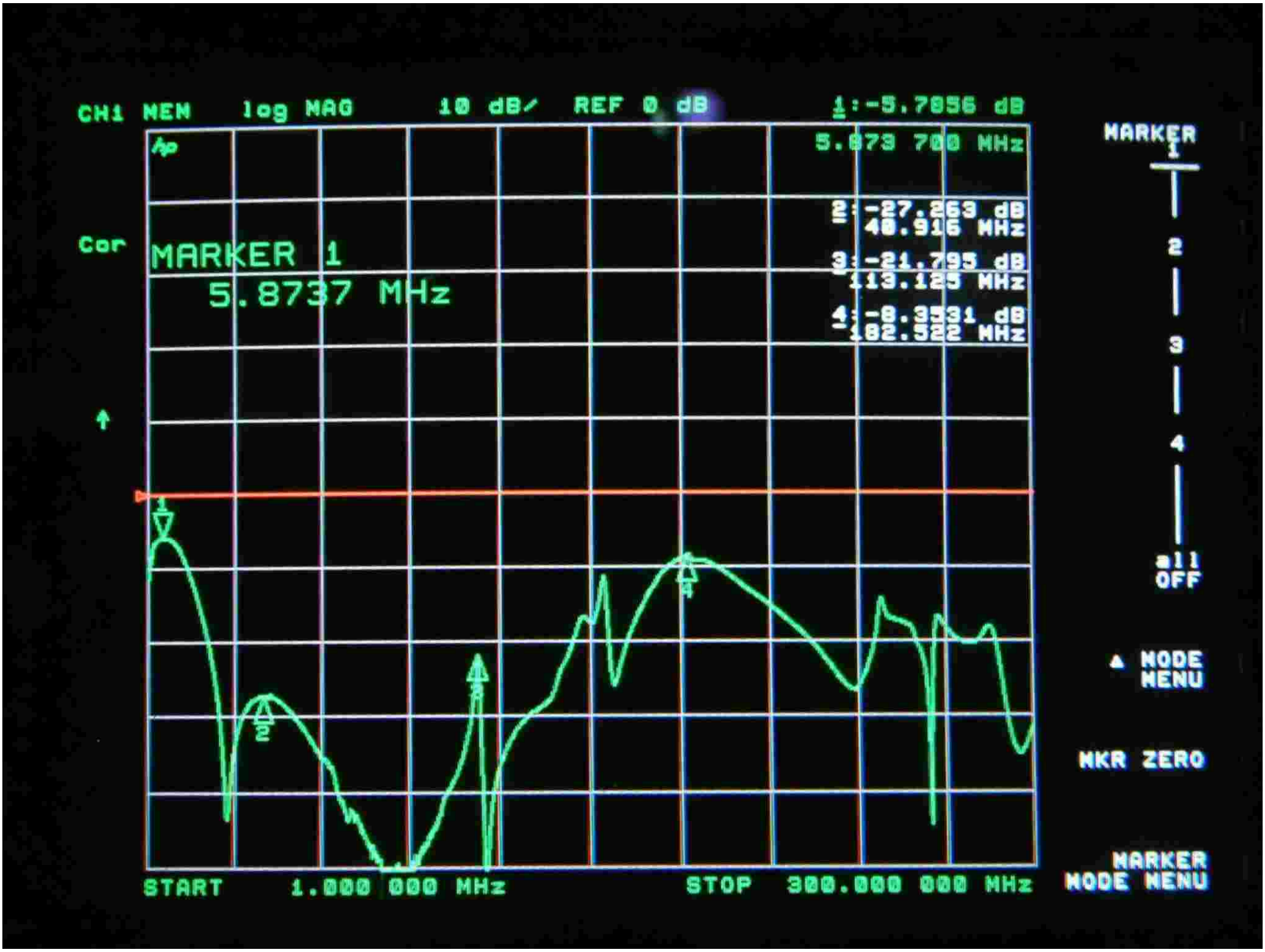

Figure 5. 15m operation. Voltage at highest VHF voltage point in system past the parasitic suppressors. This voltage was at plate blocking capacitor on plate choke side. Point 4 is 11 dB down, or 27% of anode voltage.

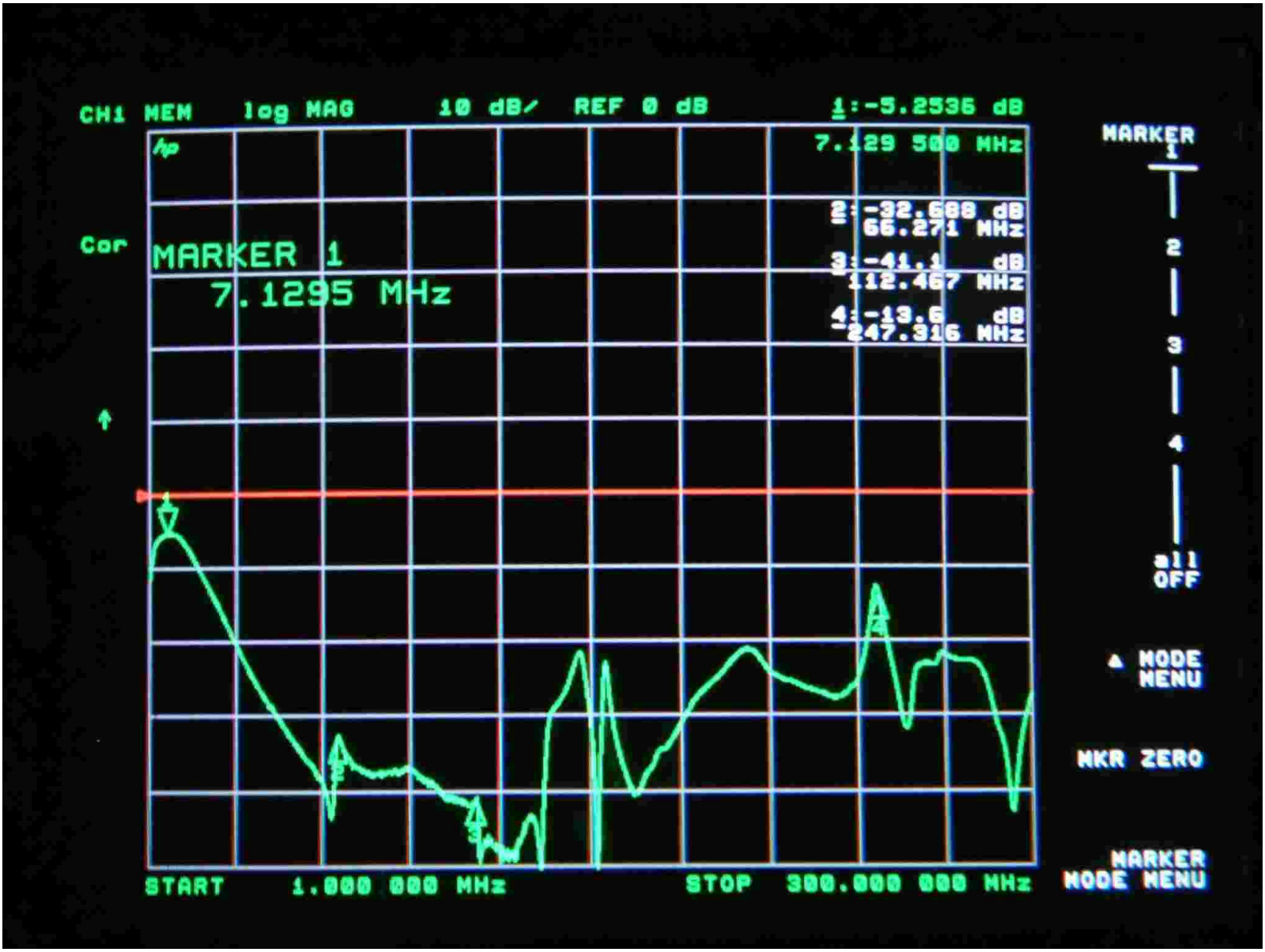

Figure 6. Voltage across loading capacitor at tank-end stud to chassis. Peak at 247MHz is -16dB, or 16% of anode voltage (480v at this point with 3000v at anode). This peak is caused by layout, and the loading cap coupling directly to tubes through stray capacitance from anode system. Impossible to have a VHF arc here, but it could aggravate TVI at USA channel 13.

Conclusion

From the above swept data, we can see VHF voltage would always be much less than HF voltage. The only likely candidate for arcing is improper operation at or near the operating frequency.

What causes arcs? Click here.

More SB221 will be added at a later date. Including performance improvement modifications.

FL2100 Problems

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The reason Svetlana tubes oscillate in the FL2100 is two-fold:

1. Bias is inadequate for Svetlana tubes. Svetlana tubes have LOWER mu (not higher as some people claim) and draw more current for a given amount of bias.
2. The FL2100 has a very poor attempt at neutralization or feedback. It also has a poor bias system.

Bias Error

The FL2100 only applies ~15 volts negative bias to the 572B grids for cutoff. This in not adequate for some brands of 572B tubes, especially those with slightly lower mu.

Feedback Error

The FL2100 design uses a capacitor from the OUTPUT of the pi network back to the input. One should NEVER do that, because the pi transforms phase along with load impedance and feedback voltage depending on control settings and load impedance!

A good design would never neutralize or add feedback from the OUTPUT of a tank circuit with a varying load back to the input.

What the design errors do in the Yaesu FL2100

When the antenna relay open, the LOAD is removed from the pi network. This INCREASES the amount of feedback by several times. The tubes, having less mu and requiring more bias, are no longer in cutoff with the relay open and draw current. This allows the tubes to amplify, and the greatly increased feedback causes the FL2100 to oscillate on whatever frequency the tank is tuned to. With no load, the voltage soars to many kV until the bandswitch arcs.

Changing tube brands in even a reasonably design amplifier is no problem at all. Changing tube brands (especially when a brand does not have the same characteristics) in an amplifier with multiple major design errors can be a big problem.

The Svetlana "fix" of increasing FL2100 bias is really not a " fix".

Corrections

Feedback

The most important "fix" would involve removing the capacitor from the output of the tank back to the tube input. This would disable the very poor feedback system.

If the amp is unstable at upper HF, a neutralization system similar to that used in the AL-811H Ameritron should be used.

Bias

The FL2100 bias system should be rewired so the relay opens the filament CT return. Normal operating bias should be left on the grids. This would FULLY cut-off any tube brand.

Switch Installation

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Switch Installation Problems

Wiring

Wiring must not have sharp points or protrusions. Solder connections should be smooth and rounded, the solder should generally for a small rounded ball that coves all sharp edges. The last motion in switch wiring to each contact should be soldering, not clipping! Clipping of protruding wires results in sharp edges. Sharp edges concentrate the electric field in a small area. This greatly decreases voltage breakdown. The difference in voltage breakdown can be profound, more than 50% reduction in voltage breakdown may result from changing a rounded connection to a sharp point.

In some cases, and intentional anti-corona ring is added near the highest-voltage contacts. Any large smooth rounded metallic object connected to the contact will disperse the electric field and reduce voltage gradient between that contact and other nearby connection or hardware. This will greatly increase switch voltage hold-off.

Contact Alignment

Switches normally have a detent located at the front of the switch. The detent causes the switch to "snap" into each band position, keeping the contacts centered.

The detent may consist of one or more balls held with spring pressure against a ball-race with recesses or dimples, or a spring-loaded cam follower that presses against a notched wheel attached to the shaft.

When the mounting nuts of supports are loosened, clearance in wafer holes and mounting brackets allows wafers to rotate a few degrees. This allows the contacts of each wafer rotor to be centered in the mating wafer stator contact. Pick-up-and-hold contacts are normally adjusted so they do not extend beyond the edge of the leading-edge stator contact.

Contact alignment is normally performed in the middle of rotation, checking by rotating in from each direction. The center position is selected, and the wafer mounting screws loosened. Each switch wafer contact group is centered or aligned, and the mounting screws retightened. After

adjustment, each position of rotation should be checked for contact centering or alignment.

Contamination

Dirt and dust are deleterious to voltage breakdown of components. They are most harmful under high humidity, when moisture and salts in the contaminant may increase conductivity. Contaminants mainly reduce voltage breakdown because they fill air gaps with slightly conductive particles.

Well-designed switches normally have much closer contact spacing than dielectric paths, because contact angles move the ends near the rotor close together. Even so, contamination of the (normally) ceramic dielectric can be an issue in conditions of severe dust buildup.

Never use contact cleaners or other lubricants on high-voltage high-current switches or relays. Many contact cleaners have carbon-based chemicals that will speed the formation of carbon tracks when corona or arcing forms. Other lubricants are very poor RF dielectrics, reducing voltage ratings.

Cleaners should be pure light hydrocarbons that fully evaporate, leaving no residual contaminants.

Preamp Summary

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When comparing low-band preamplifiers the important parameter is almost always output TOI.

Output third order intercept is a measurement standard of how well the amplifier stands up to strong signals with creating new "phantom" signals through unwanted mixing.

Type	Advertised range	NF	Gain peak	OUTPUT TOI (third order intercept)	1dB gain compression (output)
ARR 1-30	1-30MHz	5.43dB	20.4dB@13MHz	30.38dBm	21.5dBm
ARR GAsFET	single band	3.2dB	26.4dB	11.39dBm	8.1dBm
DXE RPA-1	200kHz-40MHz	3.4dB	17.3@5MHz	43.04 dBm	26dBm
ICE 124A		4.2dB		14.05 dBm	
KD9SV	160M/80M	4.9dB		14.02@full gain 0.14 dBm@12dB below full gain	

Note: *These are measurements of a single new random sample that appear to be working properly, with no noticeable defects in components or wiring.*

Vacuum Tubes

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Power Grid Tubes

Power-grid vacuum tubes cover a wide range of applications, and a wide variety of styles have evolved over the years. Every design is a compromise of some type, designed to better fit a few specific applications. For example, there are dozens of grid and cathode (electron emitter) designs. There is a wide variety of anode styles and types. Even the heaters or filaments have multitudes of shapes, materials, and physical dimensions.

The big enemy of old vacuum tubes or tubes that have sat a long time is gas. The big enemy of high gain power grid tubes (or valves) using metal oxide cathodes is grid current and excessive cathode current. Many common problems are discussed below.

Numbering System

While manufacturers use a variety of numbering systems, most Eimac tubes use an easily decoded system.

First number = number of active elements in the tube

- 2=diode 3=triode 4=tetrode 5=pentode

First letter = envelope type

- nothing or - = glass
- C = ceramic

Second letter = anode type

- nothing or - = internal
- X = external
- W = water

- V = vapor

Number group = anode dissipation (actual dissipation is much higher than number is some tubes, so check the books)

Last letter = Base or connection type

- Blank = pins or tabs
- A or Z = coaxial or pins
- F = flying leads

Last number = mu, where 1 = lowest mu through 7 = highest mu

- 1 through 5 are best suited for grid driven applications
- 7's are best for cathode driven

From this we see a 3CX3000F7 is a triode with ceramic envelope and external anode, has approximately 3000 watts dissipation (in fact it is 4000w), has flying leads, and highest mu (most suitable for grounded grid).

In addition, some tubes have additional letters inserted in the type number. Examples are the 4PR1000 or 3CPX800A7. The P generally signifies pulse rating, and the R more rugged construction.

Electron Emitters

Power-grid vacuum tubes operate by a thermionic emission process. In a vacuum tube, we commonly call the electron emitter a "cathode". The cathode can be directly heated or a filament-cathode. Another type is the indirectly heated, or heater-cathode type of electron emitter.

The cathode, regardless of sub-category, uses a material easily able to "give-up" electrons with heat. The cathode is heated in a nearly perfect vacuum to temperatures above 1000K (we would see that as a red glow) to as 2600K (a yellowish-white glow). This heat supplies energy to loosely bound electrons at the cathodes surface, causing some electrons to escape the material. These electrons form a cloud suspended around the cathode.

Commonly Used Cathode Materials

Material	Operating Temp Range K	Emission mA/watt	Emission mA per cm^2	Special Features

Pure Tungsten	2400-2600	2-10	100-1000	Immune to positive ions, can be operated at highest anode voltages. Instant on.
Thoriated Tungsten	1800-2000	50-100	700-3000	Much higher emission, but less immune to positive ions. Can be operated at reduced filament voltage or operated with peak currents over emission limits with little effect on life. Instant on.
Oxide-filaments	1000-1100	200-1000	400-3000	Very high emission per watt of filament power, but extremely sensitive to positive ions. Emission failures occur from low temperature operation if electron cloud is depleted. Must be thoroughly warmed up before any current is drawn, but warm-up time is short.
Oxide-cathodes	1000-1100	10-200	1000-3000	High emission in small cathode. Long warm-up time. High peak emission for pulse applications.

Positive ions are formed through ionization of residual gasses. High anode voltages speed formation of positive ions. The space charge (electron cloud) surrounding the cathode repels positive ions, and keeps the ions from poisoning the cathode. When positive ions contaminate the cathode, electron emission is reduced.

Tungsten has the highest melting temperature of any metal, above 3600K. Tungsten has the ability to operate at high temperatures for many hours without evaporation of materials becoming a major problem. Tungsten is one of the few materials able to withstand total depletion of the electron cloud in a vacuum tube. Tungsten emitters are not damaged by stripping away of electron clouds, and the resulting cathode bombardment by positive ions. This immunity to damage means emission life is

not shortened by excessive current, operation before full temperature is reached, or low filament voltage.

If a cathode material sensitive to positive ions is operated with excessive HV, or has excessive cathode current for the cathode operating temperature, it will suffer emission failure. This is why indirectly-heated tubes or oxide-cathode tubes must have long controlled warm-up times before any cathode current is drawn. We must ***NEVER*** operate the 3CX1500A7/ 8877, 3CX800A7, or other oxide-cathode tubes at reduced heater voltage! This would allow the cathode's protective electron cloud to be depleted, and nothing would prevent positive ions from striking and poisoning the cathode.

Anodes

The anode is operated with a positive potential. This creates a strong electric field that attracts electrons from the cathode. The electrons are accelerated (or impeded) as the pass through the opening in the grids, and eventually strike the anode at reasonably high velocity. Since the electrons have mass and speed, they have kinetic energy. This energy produces heat, as well as dislodging other electrons from the anode. The amount of heat is proportional to the velocity and number of electrons. The power converted to heat is the anode voltage times anode current, at any instant of time.

The actual plate dissipation is the time-integrated amount of anode current and voltage. In an amplifier with steady current, anode dissipation is found quite easily by multiplying the accelerating voltage (anode-cathode potential) times anode current. In an amplifier system, dissipation is a complex function of the constantly changing anode voltage and current. In such systems, a very close approximation of anode dissipation is given by multiplying average anode current by average anode voltage, and deducting the useful power extracted by the load.

Internal Anodes

Many lower power tubes use internal anodes. Internal anode tubes can use any electron emitter or grid configuration, although the most common large transmitting types use thoriated-tungsten filament/cathodes. Internal anode tubes are cooled by infrared radiation, by direct thermal conduction to the envelope, and through external connections via leads exiting the envelope.

The anode often serves three very important functions in a transmitting tube:

- The anode collects most of the electron current from the electron emitter in the tube (cathode)
- The anode must dissipate heat produced from kinetic energy of electrons that strike the anode

- The anode often degasses the tube

The material used in the anode varies with tube type and manufacturer. Most transmitting tubes use molybdenum, tantalum, graphite (carbon), or other high temperature materials in the anode. The material must be mechanically and electrically stable, even while operating with very high temperatures.

In transmitting tubes, virtually all anode heating comes from the kinetic energy of electrons striking the anode. The actual resistance of materials in the anode (and other elements in the tube) is very low, and the relatively small amount of $I^2 \cdot R$ heating is dwarfed by the hat caused by electrons impacting the surface of materials in the tube.

Higher power internal anode tubes almost always have a gettering material coated directly on the anode. The getter acts as a sponge, soaking up any gas molecules inside the tube. This is necessary because any gas creates positive ions, reducing tube life. Not only that, even the slightest amount of gas greatly reduces breakdown voltage through a vacuum.

Note: Peak anode voltage in a properly tuned and properly driven amplifier approaches two times the dc anode potential, and under conditions of mistuning, load faults, or excessive drive anode voltage can be several times the dc anode potential. It is important that any tube have an anode breakdown voltage at least four or more times the expected highest dc anode potential, or occasional anode to grid (and cathode) arcing may occur.

The most common gettering material is zirconium. For example, zirconium is used on the outside of graphite or molybdenum anodes in 3-500Z and other tubes. It is the dull gray powdery or grainy texture coating you see on the surface of 3-500Z, 811A, and 572B anodes.

Zirconium getters best at about 1000 degrees C, this is why large metal anode transmitting tubes like the 4-400A, 4-1000A, and 3-500Z must be operated with a dull red to red anode color. Zirconium also releases some gasses and absorbs other gasses at various temperatures. The varying temperature across the length of the anode (and as the anode heats and cools) allows the gettering agent to absorb a wide variety of gasses.

The quickest way to ruin a 3-500Z, or other glass power grid tube, is to never show anode color over a prolonged period of time! Storing a 3-500Z for many years without operation almost guarantees a flash-over will occur at the first application of high peak anode voltage.

Gas, either from poor manufacturing processing, out-gassing from elements, or seal leakage is the primary failure mode of glass tubes. Secondary to high-vacuum (gas) arcs are problems like bad welds or grid or cathode materials that fall out of alignment due to thermal or mechanical stresses.

External Anodes

We commonly associate external anodes with expensive tubes having fragile low-dissipation grids, long warm-up time oxide cathodes, and poorer reliability of tubes like the 3CX1500A7/8877, 8874, or 3CPX5000A7.

External anode tubes, just like *internal* anode tubes, can use any type of grid and cathode structure.

3CX1200's, 3CX3000's, and 3CX10,000's are popular tubes using thoriated-tungsten emitters. They have very long life, nearly instant warm-up, and very rugged grids.

External anode construction provides three primary advantages, all of which center around size reduction:

- More dissipation (power) can be handled in a smaller package
- Compact size allows better operation at higher frequencies
- Gain is higher, because electrons are more focused and better controlled by the electrostatic field of grids

Only oxide-cathode tubes enjoy the last two advantages. Oxide cathodes provide a very compact high-current emitter. Normally cathode oxide is deposited in bands or rings, and grid wires are aligned directly over the gaps in oxide bands. The 8877, for example, has fairly large diameter cathode (over 1"). More than 100 concentric bands of oxide are deposited on the cathode, and a grid wire is aligned just outside the area of each emission band.

This construction allows the control grid's electrostatic field to control emission, yet keep grid wires away from the primary cathode-to-anode electron streams. The grid is placed very close to the cathode (thousandth's of an inch), while the anode is much further away.

The result is very low grid current (grid intercept) from wire placement out of the electron stream, and very high gain from the very high ratio of electric fields (from distant anode and very close grid) reaching the cathode electron cloud.

Unfortunately the very same things that create very high gain also cause manufacturing and potential operating problems. The close spacing of cathode and grid increases the chance of grid-cathode shorts. Critical placement of grid wires outside of the many cathode bands makes the grid susceptible to alignment problems. If every grid wire, out of hundreds of wires, is not perfectly aligned outside the electron stream electrons will impact one or more wires and cause gold to migrate off the grid. The mechanism is much like water evaporation, rather than a catastrophic event like boiling or melting. The gold will slowly evaporate and redeposit elsewhere in the tube, either causing arcs, shorts, or poisoning of cathode oxide. This is the most common failure of metal-

oxide cathode tubes.

Ceramic tubes with thoriated tungsten electron emitters have much longer life and higher reliability than any other type of power grid tube, as a general rule. They combine the best of both worlds, having the low seal leakage and low out-gassing typical of ceramic external anode tubes and the rugged wider-spaced high dissipation grids.

The only disadvantage of thoriated-tungsten ceramic tubes is gain and frequency response is generally less than provided by oxide-cathode tubes.

The most common failures in thoriated-tungsten ceramic tubes relate to old age, and loss of emission. Ceramic tubes have very good shelf life, unlike large glass tubes.

Grids

Grid control the movement of electrons inside the tube through electric fields surrounding the grid wires. The electric field extends beyond the many wires making up the grid, almost completely dominating the high electric field created by the anode.

The anode and grid normally operate at cooler temperatures than those where robust thermionic emission of electrons occur. With little or no thermionic emission, the electric potential difference (strong electric field gradient) between the anode and grid(s) results in minimal current flow (electron movement) between anode and grid. Most of the current that does flow between anode and grid occurs because of stay gas molecules creating ions or electrons dislodged from the anode by kinetic energy of cathode-to-anode electron flow as electrons impact the anode.

Elevated grid temperatures, primarily caused by kinetic energy of cathode electrons striking the grid, may eventually reach magnitudes where the grid actually starts to show thermionic emission. Many people assume the grid dissipate rating of a tube is a value set by grid failure, but that is incorrect. The dissipation rating is actually a power related value, indicating the power required to heat the grid enough to start thermionic emission.

Eimac determines rated grid dissipation of a tube by increasing long-term grid dissipation while periodically checking at very short intervals for thermionic emission from the grid.

In actuality, tungsten grids in tubes like 3-500Z's can be operated at temperatures where they show color without permanent damage. While tube operation at such temperatures is compromised, the grid and tube often suffers no permanent damage.

The same is not true for gold-plated grids, like those in metal-oxide cathode tubes. Grids plated with soft poorly-bound materials like gold suffer gradual long-term deterioration even with

moderately low levels of grid current. Heat the grid of a metal-oxide cathode tube to temperatures even well below incandescence, and the result is instant irreversible failure.

This is why ALL metal-oxide cathode tubes should have a fast-acting electronic grid protection systems, and why fuses and worse yet resistors intended as fuses offer no protection at all to grids in metal-oxide cathode amplifiers.

Just as anode dissipation can not be determined by simply multiplying Ip times Ep in a working amplifier, control grid dissipation can not be determined by simply multiplying voltage by current in a sub-class 2 amplifier (i.e. AB2). Beware of any article or author who tells you RF grid voltage can simply be multiplied by average grid current to determine grid dissipation.

Grid dissipation, like anode dissipation, is a time-integrated function of instantaneous dissipation throughout the RF cycle. Computer models have made the difficult task of calculating grid current simple and accurate. Short of that you need to do an actual complex analysis of the grid system, such as in a Chaffee analysis.

In grid driven stages, assuming total power applied at the grid is accurately known, grid dissipation is determined by deducting bias power from RF grid power. This is very similar to the short-form determination of anode dissipation from anode power and RF power output. This does not easily work in cathode-driven PA's, because drive power is mixed with output power.

Grids are normally supported only at one or two places, and use very small wires. Normally the base material forming the grid is selected to be very hard, and to have minimal movement over wide temperature changes. The normal base metal in high power tubes is tungsten, just as used in filaments. Metal-oxide tubes require a gold overlay, to prevent contamination of materials.

Grid failures occur from four primary causes:

- Poor alignment of the grid causes some areas to "hot spot" without overall current being high
- Excessive grid current in metal-oxide tubes causes gold to migrate
- Gold plating is not always bonded properly, resulting in sputtering or flaking of gold
- Poor welds or material problems result in grids warping or moving into the cathode

Non tube-manufacturer related grid failures, like anode failures, are generally the result of a series of somewhat lengthy abuse spread over long periods of time.

Using Old Tubes

Tubes that have "set up" for a while often collect gas. This gas either comes from slow leakage through tube seals or outgassing of tube elements. This is the number one problem with old vacuum

tubes. The section below describes how to remove gas in old tubes.

Gettering and Arcing

Gettering is very important, since even miniscule amounts of gas will cause a low-resistance arc from anode to grid or cathode. The normal results of such arcs are blown grid chokes, collapsed anode chokes, damaged meter shunts, and other problems cause by high fault currents. While a few people blame high fault currents on parasitics, it is actually impossible for a parasitic to create such arcs. All the anode and grid can do is deplete the electron cloud from the area of the cathode, and the available current even with a parasitic is limited by the available emission. Uncontrolled arcs are always the result of gas or element alignment in the tube, rather than excessive current from oscillations.

Anode systems should have series-resistance to limit peak current in the event of a tube arc or failure. That series resistance should always be in the anode lead between the filter capacitors and RF choke. A diode clamp should be installed to protect meters, especially the grid meter since the grid is in the normal path of any internal tube arc.

Ceramic tubes, because of low anode operating temperatures, have the gettering agent applied to the cathode or filament assembly. This is the only area inside the tube that heats enough to activate most common gettering materials. **Ceramic tubes without internal flaws or broken seals can generally be gettered by running the filament at rated voltage for an extended period of time before application of any high voltage. The normal time for gettering is between one hour and one full day. If the tube does not getter within a day it is most likely never going to be restored to a operational relatively pure vacuum.**

Glass internal anode tubes generally have the gettering material coated on anodes, which must be operated a high temperatures to activate the getter. Glass tubes have a propensity for seal leakage and element out-gassing, both of which lead to a short self life for large tubes.

Under some conditions a glass tube can be restored to operation by running low anode voltages and positive bias on the grid. This will sometimes allow full operating anode temperatures to be reached, and the tube can be "cooked" for several hours. I've had about a 50% success rate restoring old 3-500Z's that have sat for years without use, and that arced when started at full voltage, by cooking them this way.

Most important, an arc by itself will break down and getter gas inside the tube. This is why an amplifier with a gassy tube will sometimes operate without problem after a sudden tube arc.

Intentional arcing and overheating, while pumping down a tube, are often a normal part of tube manufacturing processes.

Filaments

Filament life in vacuum tubes, like filament life in light bulbs is a complex function of operating hours, hot and cold cycles (thermal shock), mechanical shock, and temperature.

Other than avoiding physical shocks to the tube, or cycling the filament off and on needlessly, temperature is the only variable we can control. We should be sure the filament is operated at or below the maximum recommended voltage.

In amateur service, filaments and heaters rarely (if ever) are operated enough hours to have operating-time related failures. Most failures are mechanical failures, more related to constant on-and-off cycling of the tube rather than wear.

Mechanical issues such as material or assembly defects can combine with close element spacings inside the tube, and cause failures. With spacings in thousands of inches, even the slightest change in physical shape or positioning filaments or cathodes can create a grid-to-filament (or cathode) shorts. Material quality and manufacturing techniques by tube manufacturers are critical to tube life, and sometimes mistakes are made.

Some claims are made that oscillations can cause filaments to bend, but there isn't anything that even remotely supports such claims.

For example, the normal peak filament current of a sine-wave powered 3-500Z filament is about 1.414 times 15-amperes, or 21 amperes. The total force on the filament helice is 11 grams distributed over a 7 cm area for this 21-ampere current in the typical helical structural dimensions of the filament.

The small additional force of a fully saturated emission condition (which requires around 1000 volts of positive grid voltage) results in only 30% increase in total bending force, hardly significant in any filament robust enough to last years in normal operation. Not only can the peak current not reach fatal levels though any type of emission, the grid-cathode potential is not able to reach levels required to saturate filaments.

Why Does The Tube Go Bang?

In order for any tube under any condition to draw more than the peak saturated current, operating potential has to exceed the breakdown voltage of the tube.

There are two reasons excessive current might flow between elements in a tube:

- The peak anode supply voltage exceeds the hold-off voltage of the tube, causing a good tube

to arc
- The tube may be defective either through gas or an incorrectly located anode or grid

Either of these conditions would allow an internal arc, with fault current limited mostly by external circuit resistances. Oscillations, like normal drive power, can only take a healthy tube up to the saturated emission limits. Even then it would take hundreds or thousands of grid-cathode volts to saturate the emission, an impossible condition unless someone drives a 3-500Z mistakenly with a 4CX5000 or some other large PA!

There is a special condition where tubes can fail, and that is where excessively high voltage components are used in tank systems. Under load fault conditions, when nothing absorbs energy supplied to the tank and tank components do not saturate or breakdown, the tube peak anode voltage can reach levels where an arc occurs in a healthy tube. See the load faults and tuning section, and the practical demonstration areas of this website.

Summary

Tubes are relatively robust components, but they are one of the primary points of failure in any electronic device that uses tubes. While many of us would like to believe we can make small circuit changes to prolong tube life, the fact is most tube failures are related to material or construction problems within the tube. In amateur service, most failures relate to manufacturing problems in the tube.

Out-gassing and seal leakage are particularly troublesome in glass tubes, since porous anodes and imperfect glass-to-metal seals allow miniscule amounts of gas to enter the vacuum.

To insure maximum life in Amateur Service using thoriated tungsten tubes:

- Keep seals below rated temperature
- Avoid long periods of excessive anode or grid dissipation
- Avoid extended periods of non-operation
- Avoid filament voltage OVER the manufacturer's rating
- Do not excessively cycle the filament
- Do not subject the tube to shock or vibration
- Use some form of reliable fault-current limiting in the anode

To insure maximum life in Amateur Service using metal-oxide cathode tubes:

- Keep seals below rated temperature
- Avoid long periods of excessive anode dissipation
- Avoid even very short periods of excessive grid dissipation by including a fast electronic

grid-trip circuit

- Avoid filament voltage OVER or UNDER the manufacturer's rating
- Do not excessively cycle the filament
- Always allow full warm-up before allowing any cathode current to flow
- Never apply excessive high voltage, as it may strip or poison the cathode
- Do not subject the tube to shock or vibration
- Use some form of reliable fault-current limiting in the anode

Vacuum Tube Amplifiers

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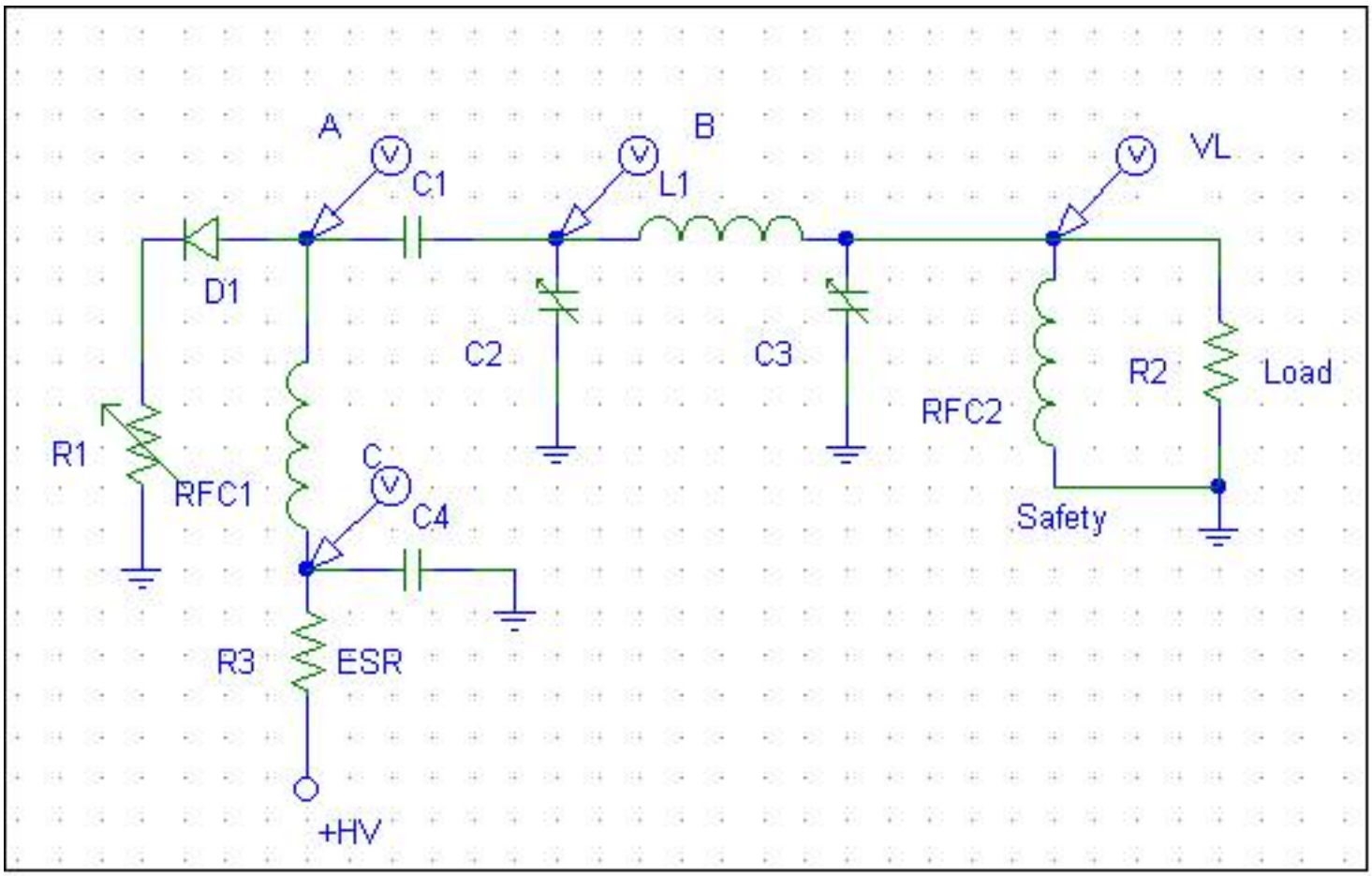
How the Tube Converts DC to RF Power

A typical vacuum tube radio-frequency amplifier has a high voltage power source. This power source supplies the energy for the RF output power. The vacuum tube acts like a non-linear variable resistance in series with a diode. The electrical equivalent of the anode and output system is:

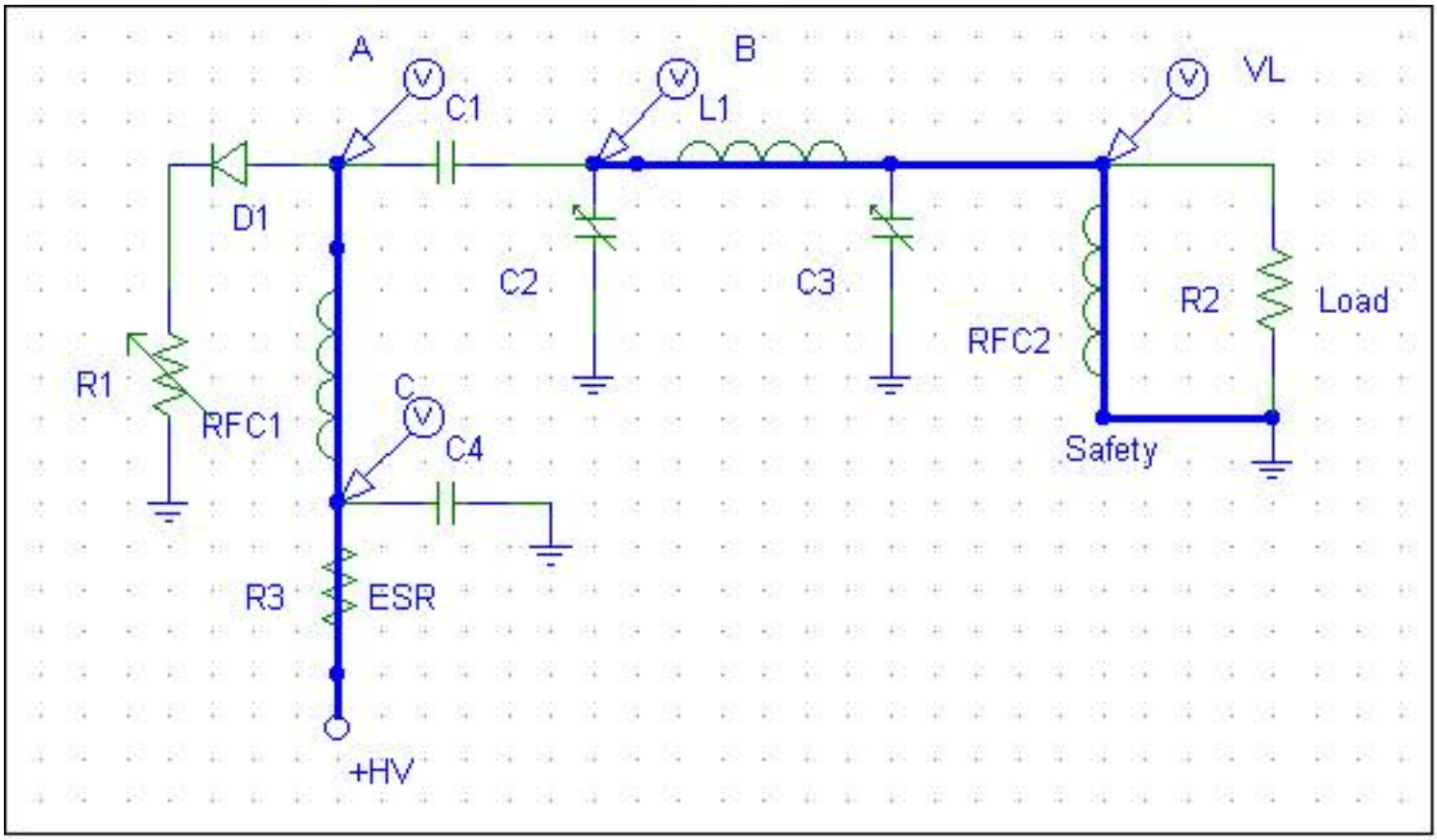


R1 represents the time-varying anode to cathode resistance, as grid-cathode and anode-cathode voltage changes, R1's value changes.

D1 represents the directional characteristics of the tube anode-to-cathode path. It is always “on” when the voltage at “A” is more positive than ground. D1 would also behave as a current limiter, the limit being somewhat less than the saturated emission current of the cathode.

Let’s examine how this circuit converts DC to RF power. We will assume +HV is 3000 volts.

In the initial state, C1 has been charged through the following DC path:



Hopefully the designer was smart enough to include RFC2, to prevent charging or discharging C1 into the load, which might be a sensitive piece of gear or the operator!

The voltage across C1 is the anode-to-ground voltage at A. Everything past A, at B and VL, is zero

volts in the resting state because of the low-dc resistance of RFC2 and the nearly infinite resistance of C1. RFC1 establishes a fixed magnetic field that is set by the quiescent current through R1.

As a positive-going grid-to-cathode voltage perturbs R1, the magnetic flux in RFC1 tries to hold supply current steady. RFC1 does this by increasing the voltage across its terminals. Energy stored in the electric field of C1 tries to hold the voltage steady, and it sources any additional current the tube (R1) requires.

The anode, in effect, “tugs” or pulls point A towards ground. The initial peak current can be quite high, because C1 has a low-reactance path to ground through C2. L1 also builds up a magnetic field, as C2 develops voltage from the current flowing through C2 and C1 back to the tube. On the initial RF cycle, the voltages hardly change before the grid is moving back negative.

At a time interval equal to 1/(4F) (where F is the RF frequency), the positive grid voltage has peaked. The grid begins to swing negative (or less positive) with respect to the cathode. Anode-cathode resistance (R1) decreases.

The collapsing field in L1 (and to a much lesser extent RFC1) tries to hold current the same, and in doing so the collapsing fields supply a small amount of additional anode voltage. Since there is no forced upward swing (the tube cannot source anything), the tank components are left to pull the small voltage change back up to the original anode voltage and slightly beyond. L1 (and to a lesser extent RFC1) actually pull the anode voltage higher by an amount nearly equal to the amount the than the tube pulled the voltage down. ***The actual amount the tank pulls this voltage higher than +HV depends on the amount of energy transferred to the load compared to energy stored in the tank system!*** This is VERY important in understanding where arcs come from!

This cycle repeats over and over, and assuming the tank system at C2, L1, and C3 is resonant, the voltage at point A increases in swing while the peak current through R1 (the tube) is gradually reduced. Eventually***, in a fully loaded amplifier,*** equilibrium is reached. In a ***properly tuned amplifier*** running at maximum available power, equilibrium occurs when point “A” swings up and down an amount just under twice the anode DC voltage.

Consider a class AB 3-500Z using a 3000-volt supply. When the amplifier is tuned into the load properly at full rated power, and driven to full power, anode voltage will swing between 5500 volts maximum and 500 volts minimum at point A. This would be a ***total anode voltage swing of 5000 volts***. Peak anode current in continuous carrier operation would typically be 1.12 amperes, with a minimum R1 value of 446 ohms. This is typical of continuous CW power out of 750 watts with a single 3-500Z tube with a 190-degree conduction angle running at 400mA anode current.

Let’s assume (we actually CAN have this condition, if the amplifier and exciter relays are not sequenced correctly) we have an envelope that rises instantly. At the peak of first positive grid-cathode cycle, the anode resistance would drop to someplace well under 400 ohms (anode

resistance is non-linear with anode voltage, and is lower with higher anode voltages). The saturated anode current would easily reach 7.5 amperes, if tube emission permits.

(Emission in a directly heated thoriated-tungsten tube is typically in the range of 50-100mA per watt of heater power, large transmitting tubes being at the upper end of that range. A full-emission 3-500Z has a saturated emission current of about 7.5 amperes.)

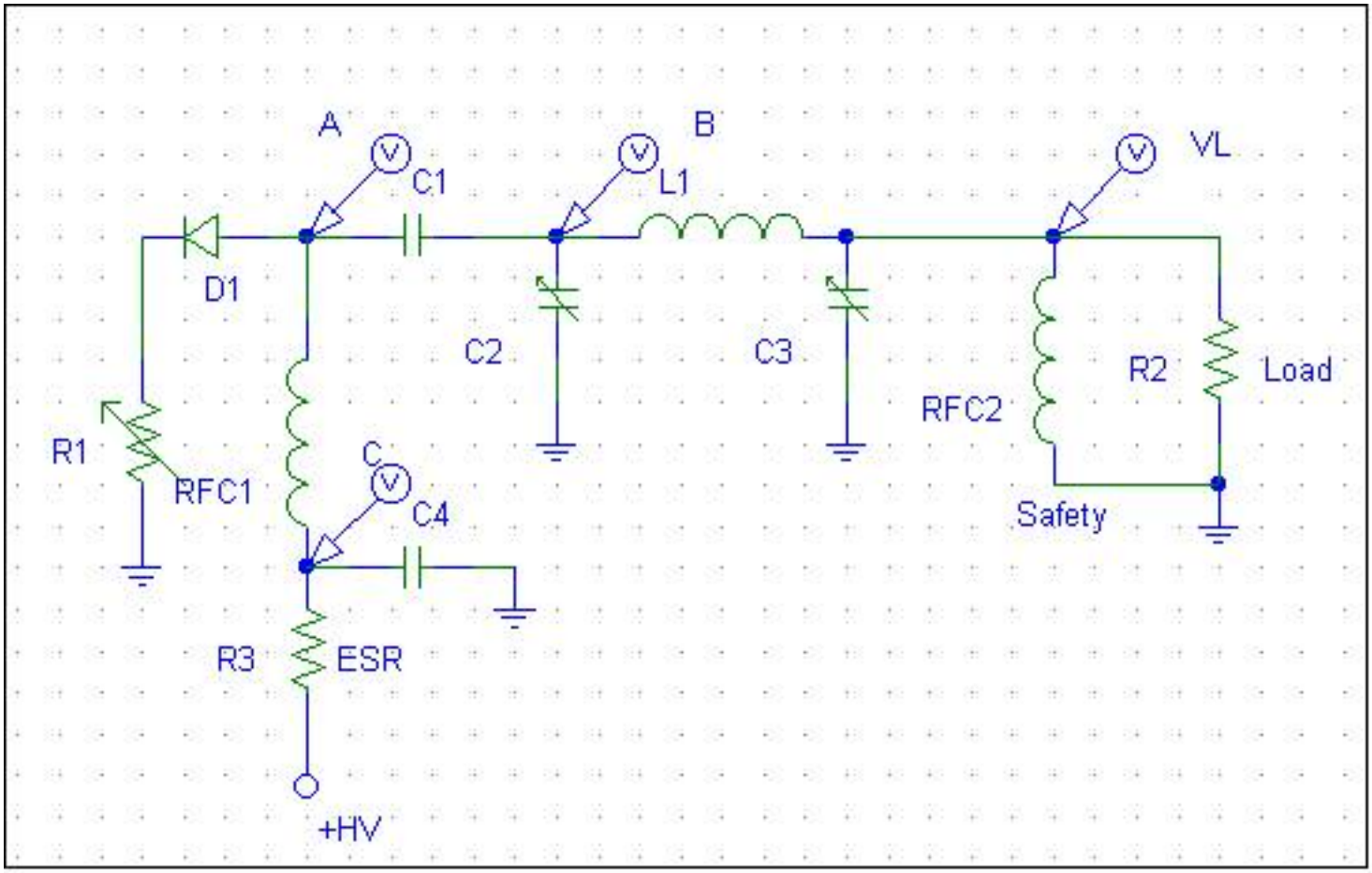
If we have infinitely fast envelope rise and fall times from the exciter, amplifier RF-envelope rise and fall times are determined by the operating Q of the tank system. C2/L1/C3 dominate the high frequency energy storage. RFC1 and C1, being larger values, dominate lower-frequency energy storage, while stored energy in the power supply dominates long-term energy demands. The voltage across C2, upon initial application of RF drive, is a rapidly expanding sine wave. It reaches maximum steady-state swing many dozens or hundreds of RF cycles later than the initial tube excitation (and decaying in a similar fashion).

The peak voltage across C2 is equal to the peak anode voltage swing, and is slightly less than twice the HV supply voltage in NORMAL operation. The voltage across C3 is a function of the load resistance, and power delivered to the load.

Incorrect Loading or Load

We know the tank circuit stores energy. We now understand the conversion process where DC is converted to AC (or RF) power. We also must understand ***energy must be transferred out of the tank at a rate equal to or exceeding the rate at which it is supplied by the downward "pull" of the tube. If we do not remove energy at a sufficient rate, voltages and currents increase until a new point of equilibrium is reached.*** Voltage at point "A" can actually swing well beyond twice +HV on upward excursions, and below zero volts (becoming negative) on downward excursions.

The maximum voltage with a load or drive fault can be tremendously higher than the typical working voltage of the tank system when energy is being removed at the proper rate.



We have the same basic tank system as discussed earlier, but *with light loading compared to drive level* the tank is pulled down very hard by the tube. Minimum Rp is reached early in the cycle, before the tank voltage reaches it's minimum swing. This is easy to do, since the tube only pulls the system down and the tank stores the energy of that downward tug.

As the plate voltage swings below zero (negative) from the tank energy, the tube is already cutting off. Nothing clamps or prevents point A from going negative. D1 effectively takes the tube out of the circuit. By the time the tank reaches it's minimum, the tube's grid-cathode voltage is already on its way positive. The tank free wheels positive, and can overshoot the +HV supply by several times the supply voltage. If loading is light enough and Q is high enough, this continues until the energy stored in the tank reaches equilibrium with energy transferred to the load, or a component fails and the arc dampens the tanks gyrations.

See the practical demonstration page.

VHF Stability

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rev PM Feb23/03

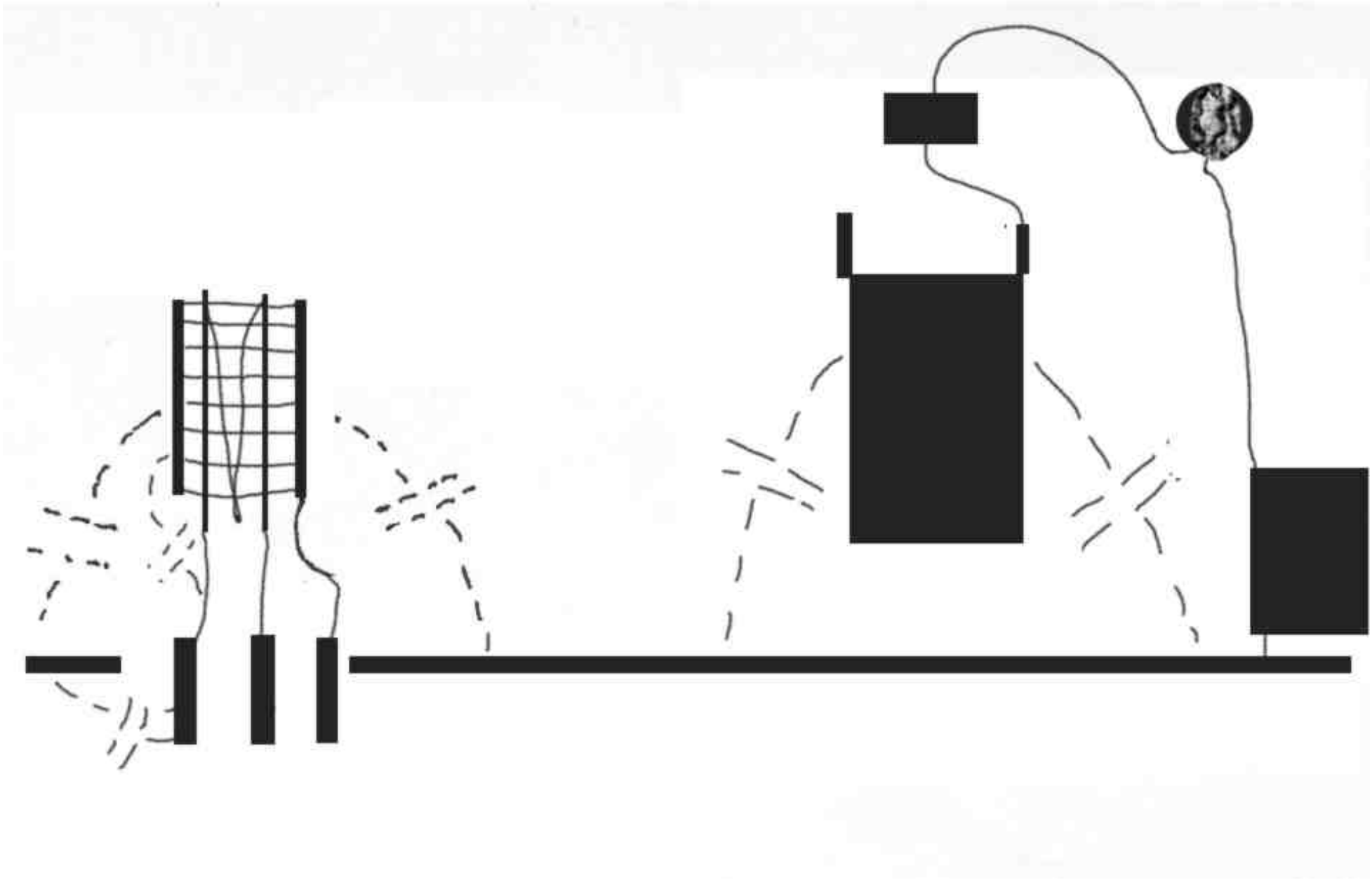
HF Amplifier Stability at VHF

A great deal of empirical engineering surrounds stabilizing amplifiers. While other forms of troubleshooting and engineering follow logical steps, it seems very few builders actually follow planned steps to test for amplifier stability. It is difficult to find detailed information describing suppressors and how they function.

Cause of VHF Oscillations

As in most planned oscillators, the most sensitive control element in the tube generally has the largest influence in determining oscillation frequency. The normal mode of VHF oscillation in HF PA's is at a frequency where the tube becomes a tuned-plate tuned grid oscillator. The control grid to anode path generally has the highest possible gain in the amplifier system, and that is why this part of the system is (by far) the most problematic area of the amplifier system.

The control grid system behaves like it is connected to a parallel tuned circuit. The stray capacitance is primarily between the grid element inside the tube and ground, generally via the filament and other connections. The inductance is via the grid leads inside the tube through the socket to the actual chassis connection. At some frequency, the grid capacitance will parallel-resonate the total inductance in the grid to ground path.



The anode is the second most common problem area. If the anode also has stray capacitance to ground. The path from the anode to chassis has inductance. At some frequency, anode capacitance parallel tunes the path to ground. This resonance greatly increases anode impedance at some very high frequency.

The grid also has capacitance to the anode, and this is the feedback path.

With all of this, the circuit has everything needed to become a tuned-plate tuned-grid oscillator.

If feedback loss (attenuation) from anode-to-grid is less than tube gain at some problem frequency, the tube ***may*** oscillate. The final requirement is the phase of unwanted feedback must be a value that causes regeneration, or positive feedback. These requirements are the same in any oscillator.

Once again, the conditions required for instability are:

- Gain must exceed attenuation in the feedback path
- The grid must have a sufficiently high impedance for the amount of available feedback to cause a stability problem,
- The anode or other element involved in the oscillation process must have a sufficiently high impedance at the same frequency as the grid to cause a stability problem
- Feedback phase must be within the correct range to obtain ***positive*** feedback

If any one of these four requirements are not met, the tube will not oscillate! This is true no matter how high or low Q is in any individual path, or if the tube has suppressors or not.

Claims have been made that tubes will remain stable for years, and a "sudden event" will make the tube break into an uncontrolled oscillation. That absolutely can not happen, unless one or more of the four important system parameters above change. Once one or more parameters change, the tube will oscillate continuously until operating voltages are removed. Quite often, in fact most of the time, oscillations are not damaging. The most common effect of VHF oscillations are creation of spurious signals; ***not*** bangs, pops, or arcing bandswitches. Bangs and pops are caused by gassy tubes or other problems, while arced bandswitches (if caused by an oscillation) are generally caused by oscillations at or near the desired operating frequency!

Location of Suppressor

Suppressors are normally found in anode systems, even though outer locations will work. A VHF suppressor must be located between the tube element and a low-impedance path to ground at VHF. This is because the suppressor must be able to load or "de-Q" one or more portions of the unwanted oscillator circuit. The actual working circuit causing a VHF oscillation is almost always entirely different than what appears on the actual component-based schematic. The cathode, an element commonly involved in low-frequency instability, is rarely involved in VHF oscillations, other than supplying electrons and stray capacitance to ground.

A VHF oscillation, if it happens to occur in an HF PA, is almost always rooted in the system behaving like a "tuned-plate/tuned-grid" oscillator.

Most of our modern PA's are grounded grid (cathode driven). Cathode driven operation requires one or more grids be directly grounded to the chassis (at least for RF) with the lowest impedance possible. *This is necessary to shield the output from the input, and assure operating frequency stability and purity of emissions.*

Using the anode for suppression generally works best because the grid or grids can remain well-grounded for RF, provided the best operating frequency performance.

The Most Unstable Tubes

The most problematic tubes for VHF oscillation have relatively large elements and long thin leads. Tubes of this type have low gain or are unusable at VHF, because elements in the tube (combined with lead inductances) are actually resonant at VHF.

Leads are a particular problem. The longer and thinner the leads, the less stable a tube becomes. Long thin leads move the self-resonance lower in frequency while increasing element impedance. This allows even a tiny amount of anode-to-grid feedback to cause oscillations.

Tubes most troublesome are 811A's, 572B's, 833's, 4-1000A's, 3CX1200A7's and 3CX1200D7's.

Tubes of moderate instability are 3-500Z, 3-1000Z, and 4-400A's.

Tubes having virtually unconditional VHF stability are the 3CX800, 3CX1200Z7, 3CX1500/8877, 3CX3000, and 3CX5000/3CPX5000/YU-156 series.

Looking at the above tubes, it is the tubes with the thinnest and longest leads that are most troublesome. These also are tubes with the poorest VHF performance when used in amplifiers intended to operate at VHF.

The most troublesome tubes above tend to oscillate in the lower-VHF range, between 30 and 100 MHz. The typical instability frequency of an 811A or 572B is around 80-100MHz, assuming all leads are short.

Moderately stable tubes tend to oscillate at 100-200MHz. 3-500Z's, for example, generally are most unstable from 180-200 MHz.

Anode Circuit Layout

Anode circuit layout can contribute to VHF instability. Long thin leads from the tube anode connector to the chassis at VHF are a problem. Problems can occur when thin (and long) plate blocking capacitor leads, thin and/or long wiring, and poor mounting of the plate tuning capacitor are used. Remember, this is a VHF path also, even if the amplifier only *intentionally* operates on HF.

To maximize stability:

- Use wide anode circuit leads from the tube to the tuning capacitor
- Mount the tuning capacitor directly on the chassis, or on a large metallic groundplane area that is thoroughly bonded to the chassis at many points
- Use a low-inductance plate blocking capacitor
- Keep all leads as short as possible, even if it is at the expense of having wiring "look pretty" with all perfectly aligned 90-degree angles
- Use the chassis as a groundplane and an input/output shield, not a front panel

Grid Circuit Layout

The grid circuit layout is probably the single most important area for insuring a stable design. Long thin leads from the tube grid connector to chassis at VHF are a problem. This can be from thin (and long) grid bypass capacitors, thin and/or needlessly long wiring, and failure to ground grids directly to the chassis by mounting ground lugs directly on the chassis near the grid pins.

To maximize stability:

- Use wide low-inductance grid leads from the tube socket directly to the chassis, Connect the ground leads at the closest possible point, using ground lugs right at the grid pins (rather than using socket mounting screws) for grounding
- Use low-pass Pi-network or parallel tuned networks as input matching circuits
- Mount any swamping or load resistors near the tube, with short leads
- Mount the low-pass or bandpass input matching system near the tube, or use exceptionally low-impedance transmission lines to reach the input matching system
- Keep all grid connections as short as possible, even if it is at the expense of having wiring "look pretty" with all perfectly aligned 90-degree angles

A Bad Grid Idea

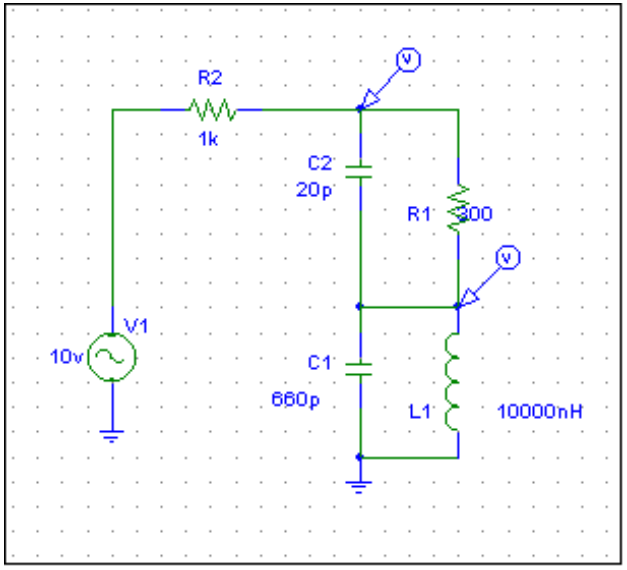
One of the very worse things in modern grounded-grid triode PA's is the inane engineering claiming floating grids on capacitors adds useful negative feedback. This is similar to what Collins did in their 811A amplifier, and Japanese manufacturers copied the bad idea into their power amplifiers. Heathkit was also a victim of this engineering gaff.

When I was designing PA's in the late 70's and early 80's, an employee of Eimac and author of many articles and a handbook put considerable pressure on me to float the grids of 3-500Z PA's through small mica capacitors.

The theory presented was pretty simple. He claimed floating grids through small mica capacitors added negative feedback, making the amplifier "work better". The basic idea was that filament-to-grid capacitance formed one part of a capacitive voltage divider, the grid-to-ground capacitors forming the other half of the divider.

The alleged "idea" was this capacitive divider would float the grid partially up from ground, and reduce grid-to-cathode (grid-to-filament) driving voltage. I quickly concluded that no one ever actually measured or calculated feedback over a wide range of operating frequencies and grid currents. Since this is a C1/C2 divider, the sampled feedback should be constant in both amplitude and phase regardless of frequency, power levels, and tuning.

The basic circuit is:



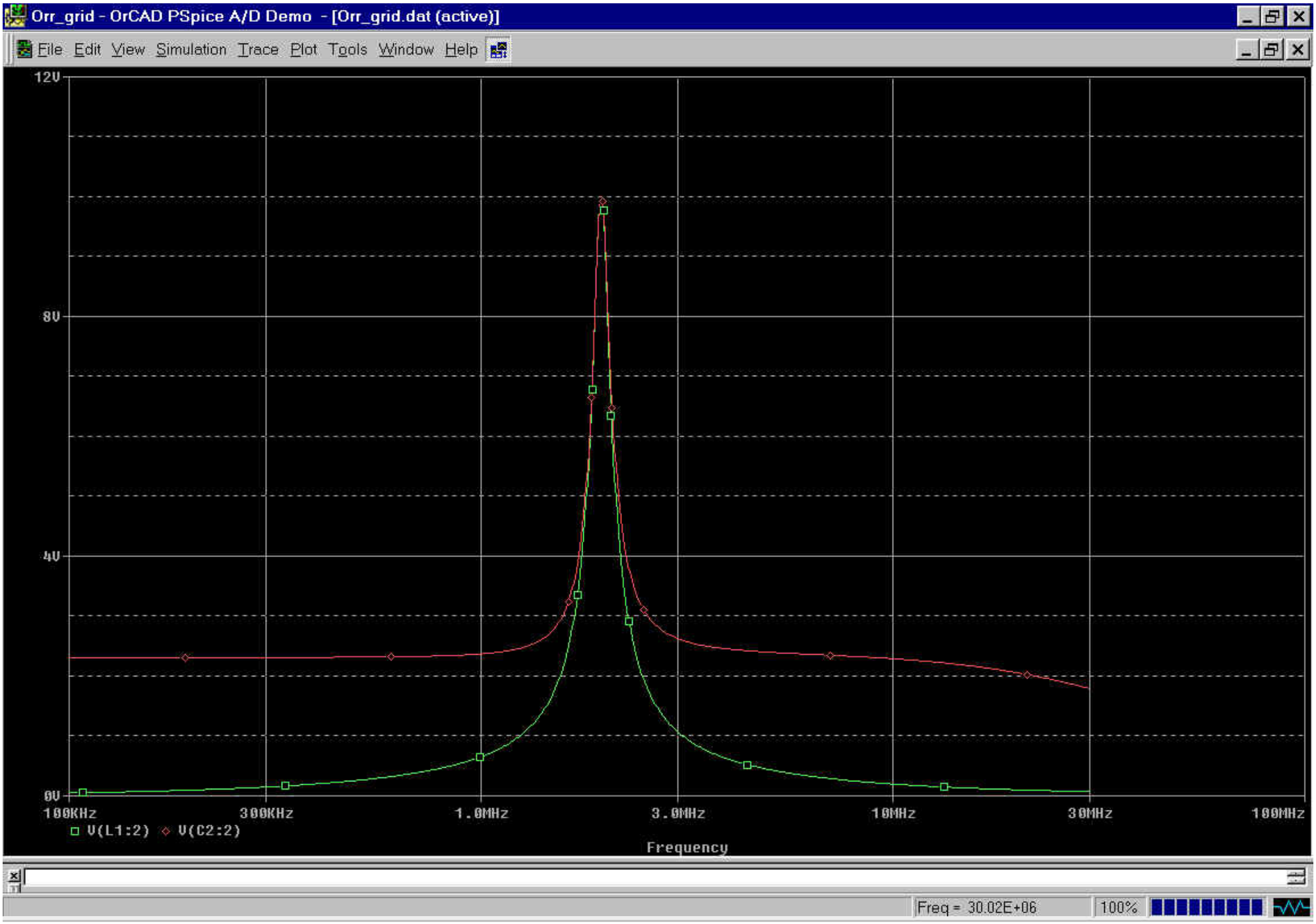
The grid connects at the junction of C1 and C2, while the cathode connects to the top of C2.

C2 is the internal stray GK capacitance of the tube

R1 is the time-varying grid impedance

R2 is added to allow us to see the input impedance change of the divider on a probe model.

Sweeping the system from 100KHz to 30MHz shows us the following:



What we see is a huge spike in grid-to-ground impedance at 2MHz, and very uneven response above that range. We did not even include the time-varying grid resistance and phase error,

since this would take up to much space on my web site.

By manipulating the value of L1 (the grid chokes) we can move the spike around, but we are ALWAYS left with some frequency where the grid isn't grounded! This is a serious violation of good engineering practices in a grounded-grid PA, and is actually at the root of stability problems in some PA's. Collins, for example, had a series of field modifications to the 30L1 grid system. The best idea would have been to abandon the bad idea that this system adds controlled negative feedback, and changed to a true grounded grid.

If they wanted negative feedback, the PROPER method would have been adding a resistor in series with the cathode by placing a series resistance immediately at the drive point to the cathode!

There are obviously major flaws with the super-cathode drive concept, when it uses a capacitor divider. Grid current causes grid-to-cathode impedance to constantly vary. When grid current is absent, the grid-to-cathode impedance is nearly an open circuit. Grid-to-cathode capacitance dominates the upper half of the divider, and everything appears to work as planned.

Unfortunately, a problem appears whenever the grid draws current. Even the tiniest amount of grid current causes grid-to-cathode impedance to decreases rapidly. With only a few dozen milliamperes of grid current, grid impedance drops to a few hundred ohms or less. As grid current is drawn, the decreasing grid impedance dominates the upper leg of the voltage division circuit!

There are also new potentially destabilizing resonances added in the grid path.

This system causes three major problems:

- Grid drive is effectively reduced as operating frequency becomes higher, just where we need the gain to flatten normal gain roll-off
- Feedback starts to show significant phase-lag with increased drive and reduced operating frequency
- Grid-to-chassis inductance at VHF and LF is increased, making the amplifier less stable

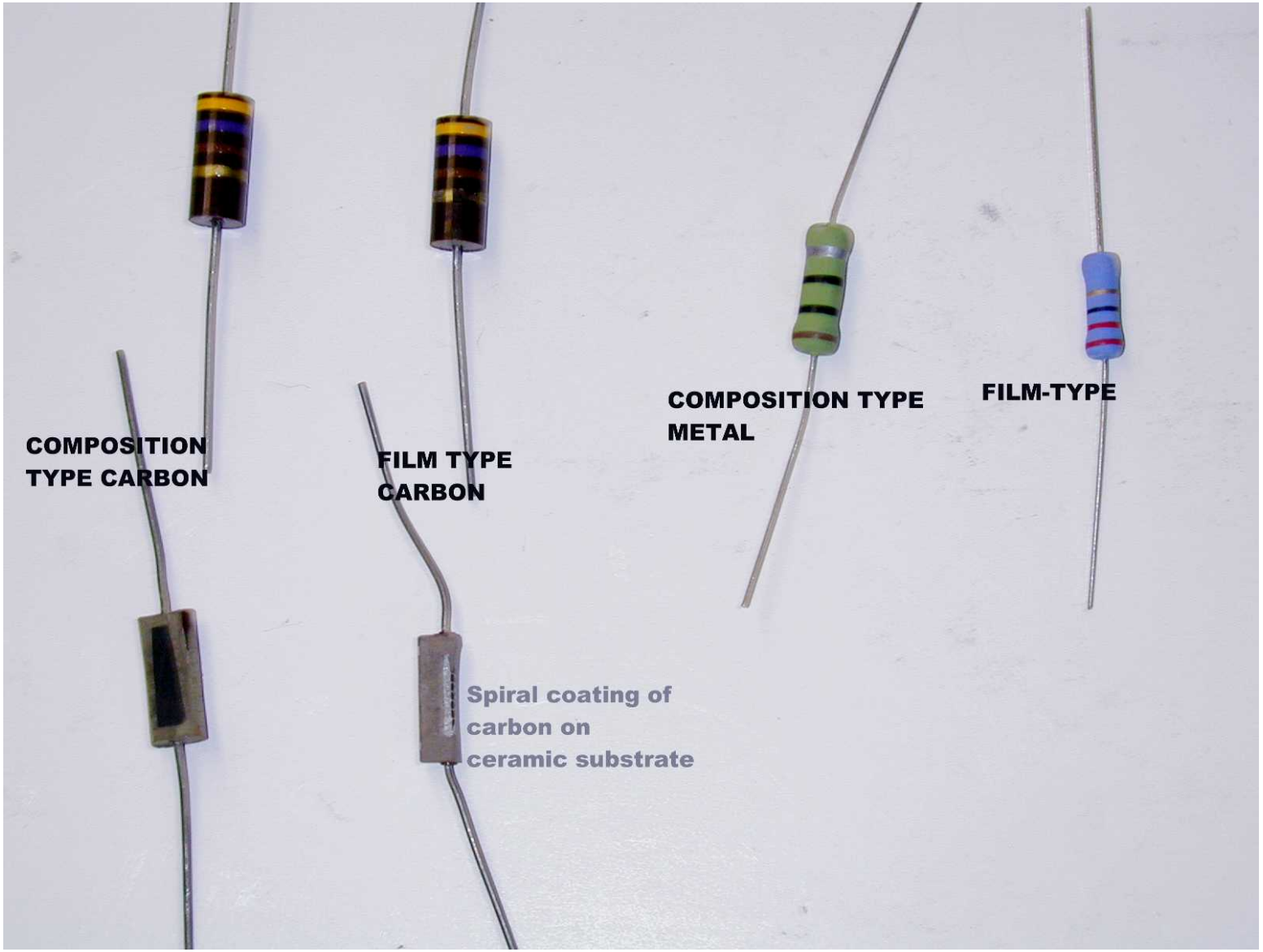
When I tested several amplifiers with and without this alleged "super-cathode" system, I found IMD performance decreased significantly under some operating conditions. Stability also significantly decreased.

Ground the grids either directly with short heavy leads or a low-inductance high-value capacitor with very ***short leads*** in any grounded-grid PA!

What Does the Parasitic Suppressor Do?

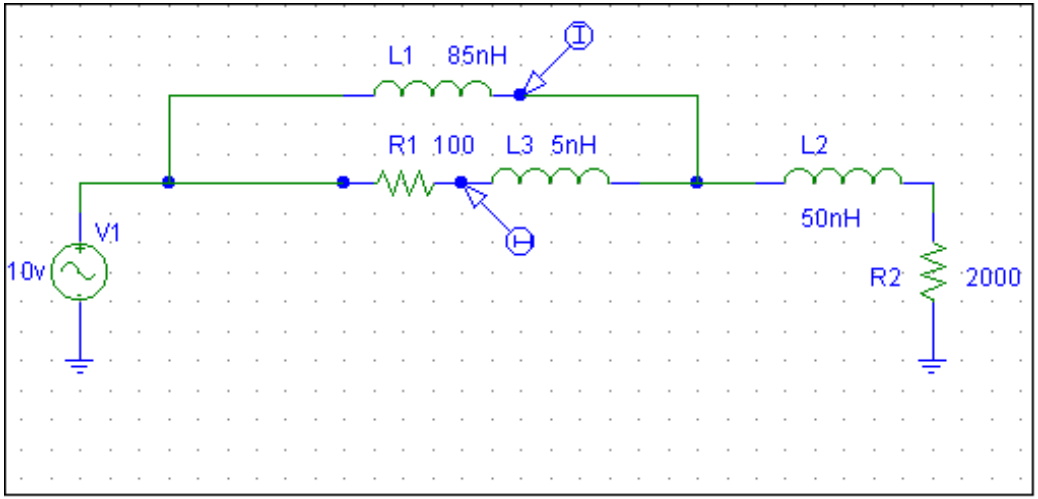
The parasitic suppressor normally has two components in parallel, a resistor and an inductor. At low frequencies, the path through the inductor dominates the system. At very high frequencies, the resistor dominates the system (assuming it is a low-inductance resistor).

One common problem is people assume brown carbon resistors are non-inductive. That isn't the case. For an example, look at the following resistors:

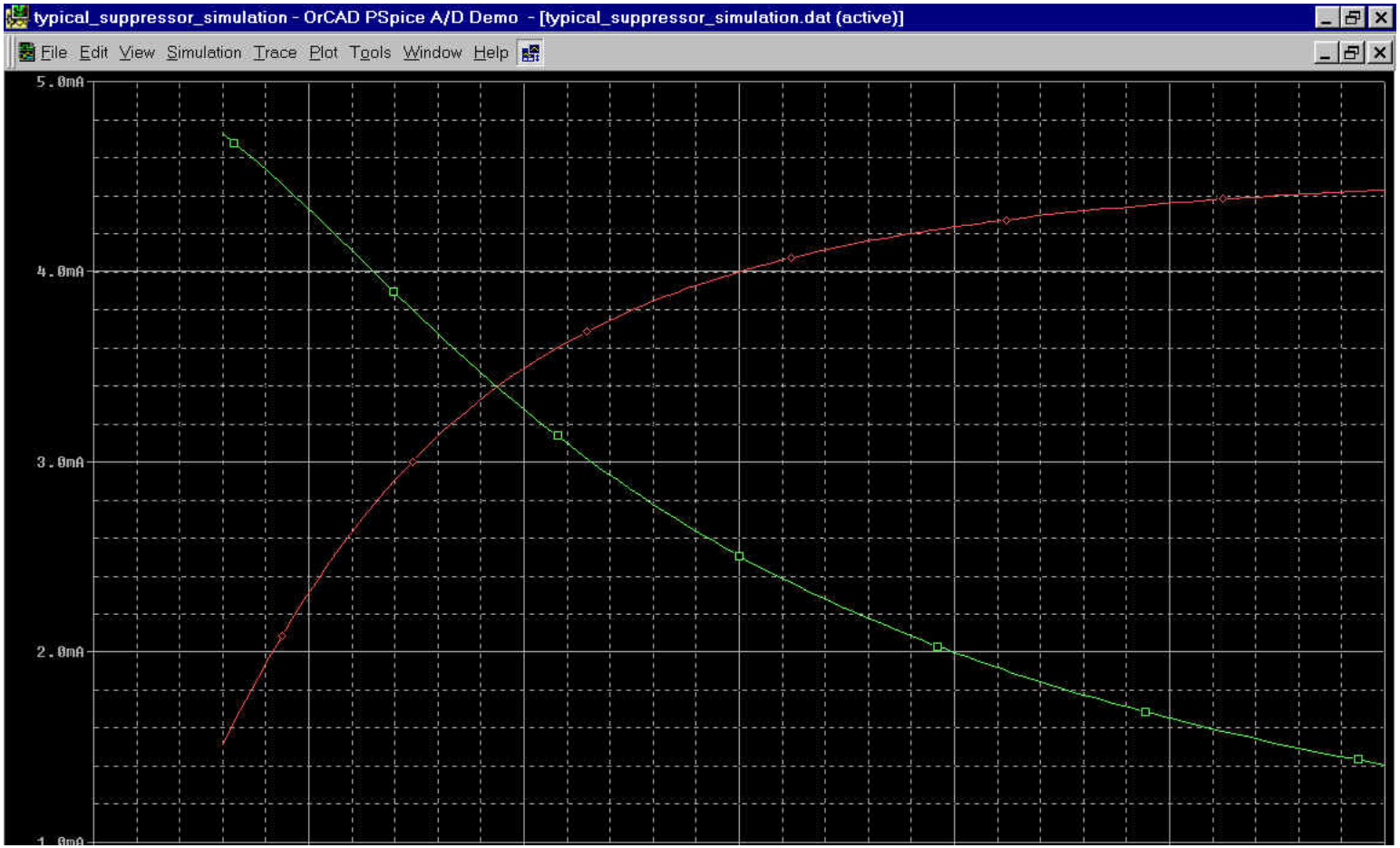


All of the spiral-conductor resistors above have significant inductance at VHF, and make very ineffective suppressors unless the reactance is cancelled. Only the true carbon composition resistors are useful in non-resonant standard suppressors.

This is a typical suppressor system, including inductance of the anode lead:



In this case V1 represents the tube. The following is a simulation of currents in the suppressor:



Starting at 30MHz, the ratio of current in the inductor to current in the resistor is:

Frequency	-I(L1)	-I(R1)
30MHz	0.0047	0.0015
60	0.0041	0.0026
90	0.0034	0.0034
120	0.0029	0.0037
160	0.0024	0.0041
190	0.0021	0.0042
220	0.0018	0.0043

This tells us something very important. The INDUCTOR dominates only at low frequencies. At 30MHz, current in the inductor is three times current in the resistor.

At 190MHz, in the range of the instability frequency of a 3-500Z, the resistor has twice the current as the inductor.

This tells us any changes in INDUCTOR design or inductor Q (such as use of nichrome wire) mainly lowers low frequency Q. It would have virtually no effect on very high frequency Q of the system.

- The dominant factor in controlling VHF Q is the resistor value, and any reactance in the resistor path
- The dominate factor in determining HF Q and performance is the inductor value, and any changes in inductor Q

This has been my point all along with the Measure's nichrome suppressor. Measures claims, incorrectly, his suppressors provide lower VHF Q while, in fact, they do exactly the opposite! A typical Measures hairpin suppressor actually produced significantly ***higher*** system Q in the anode of a 3-500Z (nearly twice the VHF Q), because the equivalent Rp of the suppressor in ***series*** with the anode lead was ***lower***!

The reasons HF PA's arc are explained at other pages of this site, and include incorrect relay sequencing, load faults, as well as improper tuning and exciter transients.

Reducing VHF Q

If we want a lower VHF Q, while maintaining high LF Q and efficiency, the system must shift current into the resistor faster as frequency increases. The suppressor must also have higher Rp, so it dominates the anode path inductance that is in series with the suppressor.

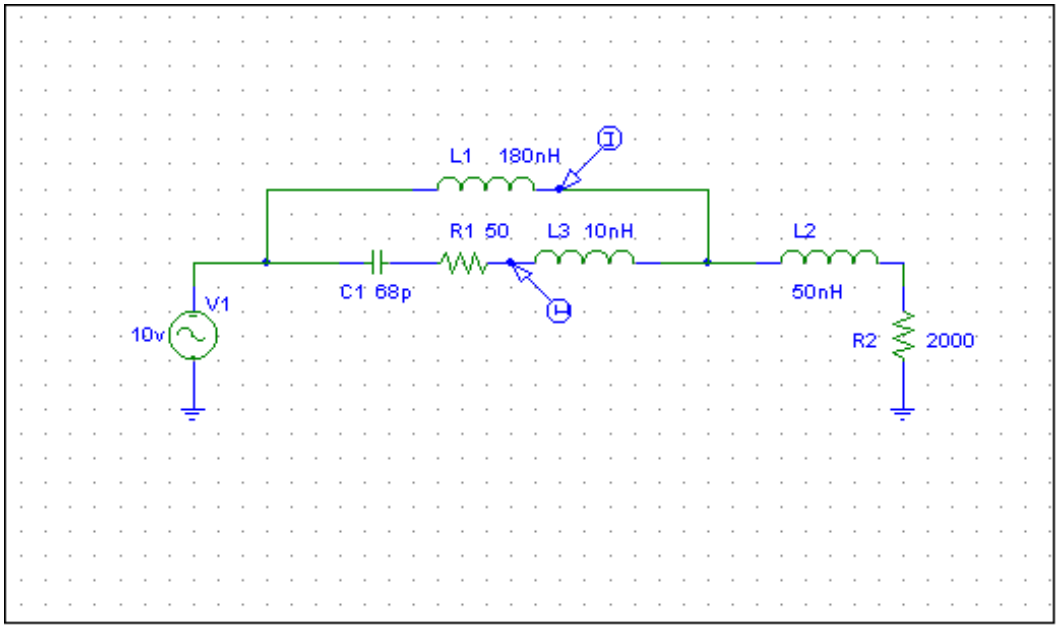
While Measures openly touts his "low-Rp suppressor", the fact is a low Rp suppressor results in higher anode system Q!

A Truly Improved Parasitic Suppressor

In order to reduce VHF Q, we must have a resistance dominate the anode system. This means, in a frequency sweep simulation, the ratio of currents in the resistance to

current in the inductance must be as high as possible. Let's call that slope ***the rate of transfer.***

The rate of transfer can be increased by adding a small value of capacitance in series with the resistor:



The old suppressor was:

Frequency	-I(L1)	-I(R1)	Ratio
30MHz	0.0047	0.0015	3
60	0.0041	0.0026	1.6
90	0.0034	0.0034	1
120	0.0029	0.0037	.78
160	0.0024	0.0041	.58
190	0.0021	0.0042	.5
220	0.0018	0.0043	.42

The new one:

Frequency	-I(L1)	-I(R1)	Ratio
30MHz	0.0069	0.0026	2.6
60	0.0050	0.0055	.9
90	0.0027	0.0052	.52
120	0.0019	0.0050	.38
160	0.0013	0.0048	.27
190	0.0011	0.0047	.23
220	0.0009	0.0047	.19

Graphically we see the currents are:

The green curve is current through the inductor, the red curve shows current through the resistor. Notice how flat current is in the resistor, and how sharp roll off of current in the inductor becomes.

This means we will have very low ***anode SYSTEM*** Q starting at a low VHF frequency of 50-60MHz, and continuing up to UHF. Dissipation in the resistor is still reasonable at HF, efficiency and tank Q at the operating frequency remain high, yet VHF suppression is greatly improved.

Selecting Component Values

Optimum resistor value can actually be determined by measurement, or determine empirically.

If the anode path is long and thin, the impedance will be high. A high anode path impedance (thin or long leads) requires higher values of resistance, because we want the resistor to dominate the anode system impedance. The best value for a resistor is generally one that is approximately equal to, or slightly higher than, the anode path reactance at the frequency of instability.

That impedance can be measured on an impedance test set, or other ways by creative engineers or technicians, but as a general rule long, thin anode leads like 811A's require 100-150 ohms of resistance while shorter thicker anode leads like those in 3-500Z tubes require 50-100 ohms of resistance. Stable tubes with external anodes often can just use anode lead resistance, using brass or other materials, to adequately dampen anode path reactance.

The inductance has to present a significantly higher reactance than the suppression resistor value at the frequency of instability. This causes the majority of current to flow through the resistance at the very high frequency, and not the inductor.

If you look at amplifier designs, you will find tubes like 811A's generally have higher value resistors and many turns of wire in the suppressor. Tubes like 3-500Z's have significantly fewer turns, especially when grid leads are kept very short and direct to the chassis, and lower value resistors.

The more unstable the amplifier tube, the larger the inductor and resistor must be.

One way to view this is to consider the frequency response of a Hi-fi amplifier. Larger values of plate load resistors in amplifier stages reduce higher-frequency gain. The same is true in HF PA's.

Lower frequencies of instability require larger inductors, so the RF path is shifted over to the resistor at a lower frequency.

Uses For Improved Suppressors

Series-resonant suppressors are used with slightly inductive resistor paths, and larger-than- normal shunt inductors. A small capacitor is placed in series with the inductive resistor path, and this capacitor series-tunes the resistor path. This results in a very rapid shift of current into the resistor as frequency is increased. This works well with amplifiers operating at 1/3 to 1/2 the instability frequency, minimizing resistor heat while providing perfect stability.

Typical applications are 3CX1200A7 and D7 tubes, 572B tubes, and 811A tubes.

Shunt suppressors with series-resonant tuning are also sometimes used, the normal application is very high power stages with substantial anode-to-tank currents. These suppressors consist of a series R/L/C system, where the C is normally just stray capacitance to the tube anode. Sometimes these suppressors take the form of a ferrite block placed between the anode and chassis. The inductance of the block series-tunes stray capacitance, and the losses act like a damping resistance in series with that path. I've stabilized 50-100kW VHF transmitter designs using shunt suppression.

Other Instability

Some PA systems are prone to oscillation at low frequencies. Yaesu and Dentron amplifiers using 572B's, and the Collins amplifier using 811A's are good examples of production amplifiers with stability problems.

These amplifiers tend to oscillate NEAR the operating frequency.

All of these amplifiers, except the Yaesu, use tubes with high anode-to-grid feed-through capacitance and no neutralization. Worse, the Collins floats the grids for RF, reducing the already poor isolation of anode-to-cathode feedback path in the 811A.

Yaesu uses one of the poorest engineered feedback systems of all, with a capacitor from the output of the pi section back to the cathode! Phase shift in that path would vary wildly with tank circuit tuning and load impedance on the PA, as would the amount of feedback!

The Yaesu amplifier is a particular problem with Chinese 572B tubes, because grid mu is lower. Negative grid bias has LESS of an effect on cathode current, so the Chinese (and Russian) tubes draw extra quiescent current when the antenna relay is open. This additional current allows the tube to amplify while the amp is in standby. Since the antenna and input source are removed in standby, and the improperly designed feedback path to the tank output remains in place, the PA oscillates near the operating frequency with no load! Voltage in the tank builds up to many thousands of volts, because no energy is extracted to a load. The fact the oscillation is at a low frequency allows the bandswitch to see the full voltage, and it fails.

Amplifiers can create extremely large voltages when RF is applied and a load is not present!

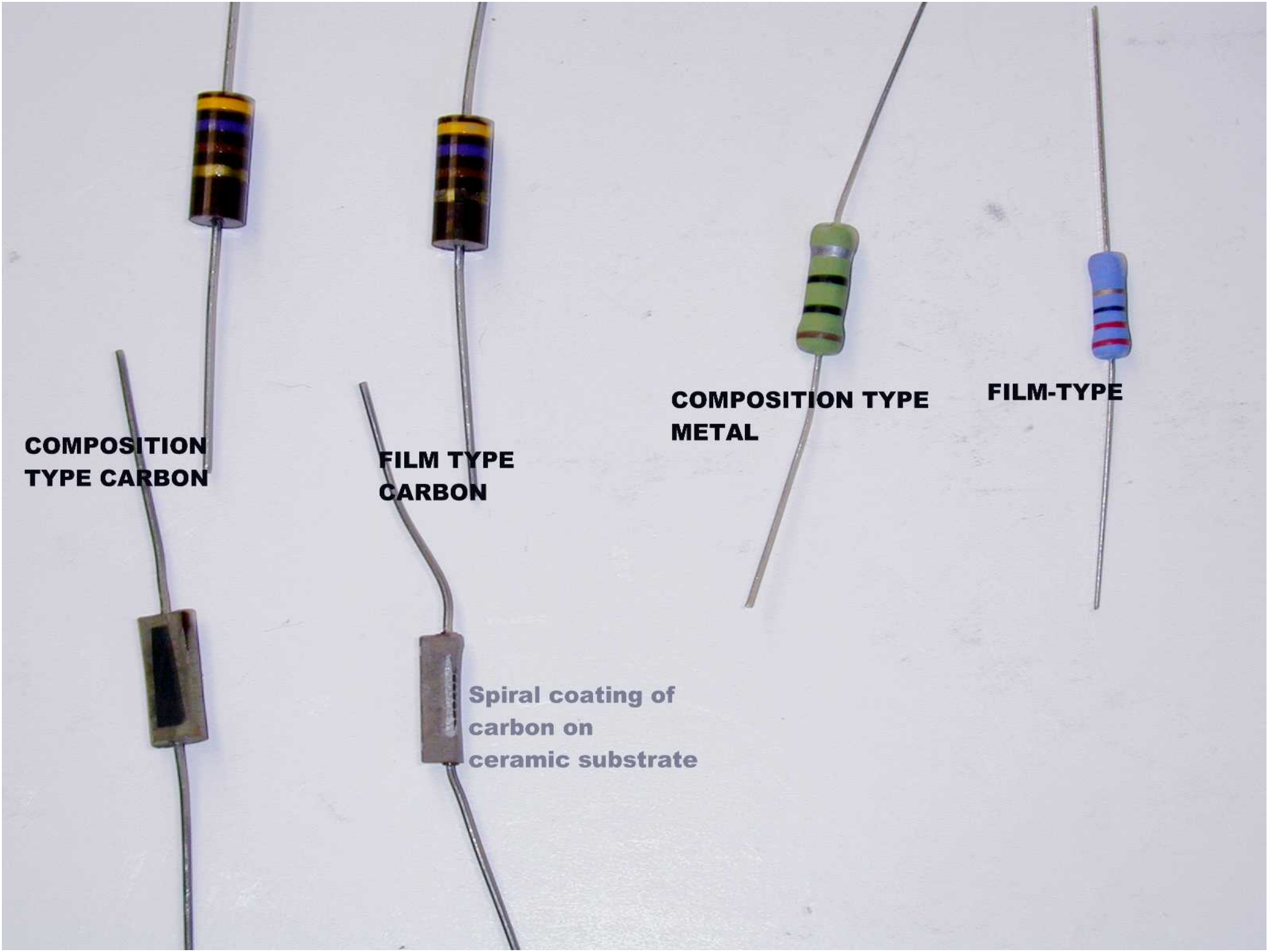
All of the amplifiers discussed above would be greatly improved by:

- Adding a proper bridge neutralization circuit like Heathkit, Ameritron, and Gonset used in 811 amplifiers.
- Grounding the grids either directly or through low reactance very-short-lead capacitors, directly between the socket's grid pin and chassis.
- Using the improved suppressor outlined above to de-Q the amp at lower VHF.

Conclusion

I hope this information is useful, and helps people understand what really goes on in a parasitic suppression system. As time permits, I add more articles about curing unique problems in amplifiers, and diagnosing amplifier failures. I hope these pages are a good start.

Please pass this web address along to others.


COMPOSITION
TYPE CARBON
FILM TYPE
CARBON
Spiral coating of
carbon on
ceramic substrate
COMPOSITION TYPE
METAL
FILM-TYPE

Sorted by IM3

[Home] [Up]

Sorted by Intermodulation. Best receiver first.

This is a measure of how badly a close-spaced signal will trash your receiving with spurious signals or splatter sounds, and you might blame it on the other guy thinking your receiver is just great. Remember that even in my very quiet rural location where I have to "beam" through the east coast to hear Europe, 80dB or so IM3 is probably more than enough. The reason my Drakes are so good is because they were built to duplex on the same band with my own transmitters running!! Unless you are in a situation like that, anything over 80dB is likely wasted.

Most transmitters are really the source of problems. The exceptions to transmitters being the primary source of problems occurs when:

1.) You use a noise blanker. Most noise blankers kill receiver IM and blocking performance.

2.) You don't have good narrow IF filters.

3.) You use too much gain in the receiver front end.

4.) You have a receiver with a design problem (less than 80-85dB close-spaced IM3 DR) .

Beyond the above, most transmitters are the real problem. Look at transmitter IM tests on SSB and read about keyclicks.

Poor Designs

Poor designs include:

1.) Many different Yaesu's without noise blanker mods. NO, changing the diodes doesn't do a thing! Been there, done that. The hissing noise comes from a FET following the narrow filters, the IM comes from the silly Yaesu noise-blanker design that hangs an FET right on a wide IF line with the potential for that FET to be at full gain even when the NB is OFF!!!

2.) DSP filter only radios. They aren't that good to date, and some of them are downright crummy.

If you have a DSP radio without a roofing filter that matches the mode, you probably have a ***poor*** receiver.

3.) Radios with spurious response problems. The JRC JST-245's I've tested and used all test good, but they have so many spurious synthesizer responses on 160 meters I'd never consider using one for serious weak signal work. I don't know how they are on other bands, but they sure stink on 160.

Here's how a few receivers stack up. Look at how much the FET in the Yaesu hurts the performance. Look at how poor a stock early R4C really is. Notice almost anything is good at 10kHz, even some very poor receivers.

	Model (in some cases serial included)	MDS	2kHz BDR	10kHz BDR	2kHz IMDR	10kHz IMDR
1	R4C heavy mod. Experimental. See note 5	-143	127	131	118	119
2	R4C 17816 med mod. See note 4	-140	96	115	92	103
3	TenTec Orion	-129	98	100	92	92
4	FT1000D/W8JI mods	-136	73	87	89	88
5	TS-870 Inrad 400Hz IF filter mod (AIP)	-139 (-125)	74 (82)	94 (98)	86 (81)	90 (91)
6	FT1000MK V W8JI mods	-131	80	89	85	90
7	IC-751A	-140	83.5	98	79	91
8	IC-7800	-127	80	98	78	88@5 kHz

9	FT1000MP 8N370165	-135	73.5	84	62* 75.5** note 1	70.5* 80** note 1
10	FT1000D	-134	68	85	72	81
11	R4C S/N 22291 stock	-139	80	116	68	86 note3
12	IC-756 PRO	-133	N/A	78	67	77
13	FT1000MK V stock	-131	74	86	60 note1A	77 note 1A
14	R4C S/N 17816 stock	-132	57	109	48	82 note2

Note 1: The FT1000MP has the same inherent problem as the FT1000 and FT1000D. **A noise-blanker amplifier FET is left on *even when the noise blanker is turned off with front panel NB and NB-W switches*.** This FET connects directly to the IF, and overloads when strong signals are present. Unlike the FT1000 and FT1000D, the front panel noise blanker level control affects the FT1000MP IMD performance. * is with the NB gain control fully advanced, ** is with the NB gain control fully off. ***In ALL cases the NB off-on switch is OFF!!***

Note 1A: Yaesu copied the same basic NB from the FT1000D, and so this radio also has a NB problem even when the NB is switched off. This test was with NB gain menu set for minimum gain. The radio is worse when NB gain is increased in the NB gain menu.

Note 2: This is a MOSFET 2nd mixer R4C, aligned and in good shape but with no changes. This receiver is almost useless for weak signal CW work of any kind because of poor IM performance and filter blow-by.

Note 3: This is a tube 2nd mixer R4C, aligned and in good shape but with no changes. This receiver is almost useless for contesting or crowded bands because of poor IM performance and bad filter blow-by.

Note 4: This is the same receiver in note 2, but with the following changes:

1. Increased screen voltage on RF amplifier to 130 volts.
2. 6J6 first mixer with oscillator injection increased to 5 volts.
3. Diode double-balanced 2nd mixer with +10dBm injection.
4. 10dB of gain removed from first mixer and first IF and redistributed behind the narrow filter.
5. Sherwood Engineering 600Hz roofing filter.
6. New AGC system. After narrow filter only, no AGC on front end.

Note 5: Gutted and rebuilt with solid-state double balanced high-level mixers. No vacuum tubes in signal path. Sherwood 600Hz roofing filter. Experimental receiver for full-duplex operation on 160 meters. This receiver allows me to hear weak DX while I am transmitting, as long as they are a few kHz off frequency.

Detuning Towers

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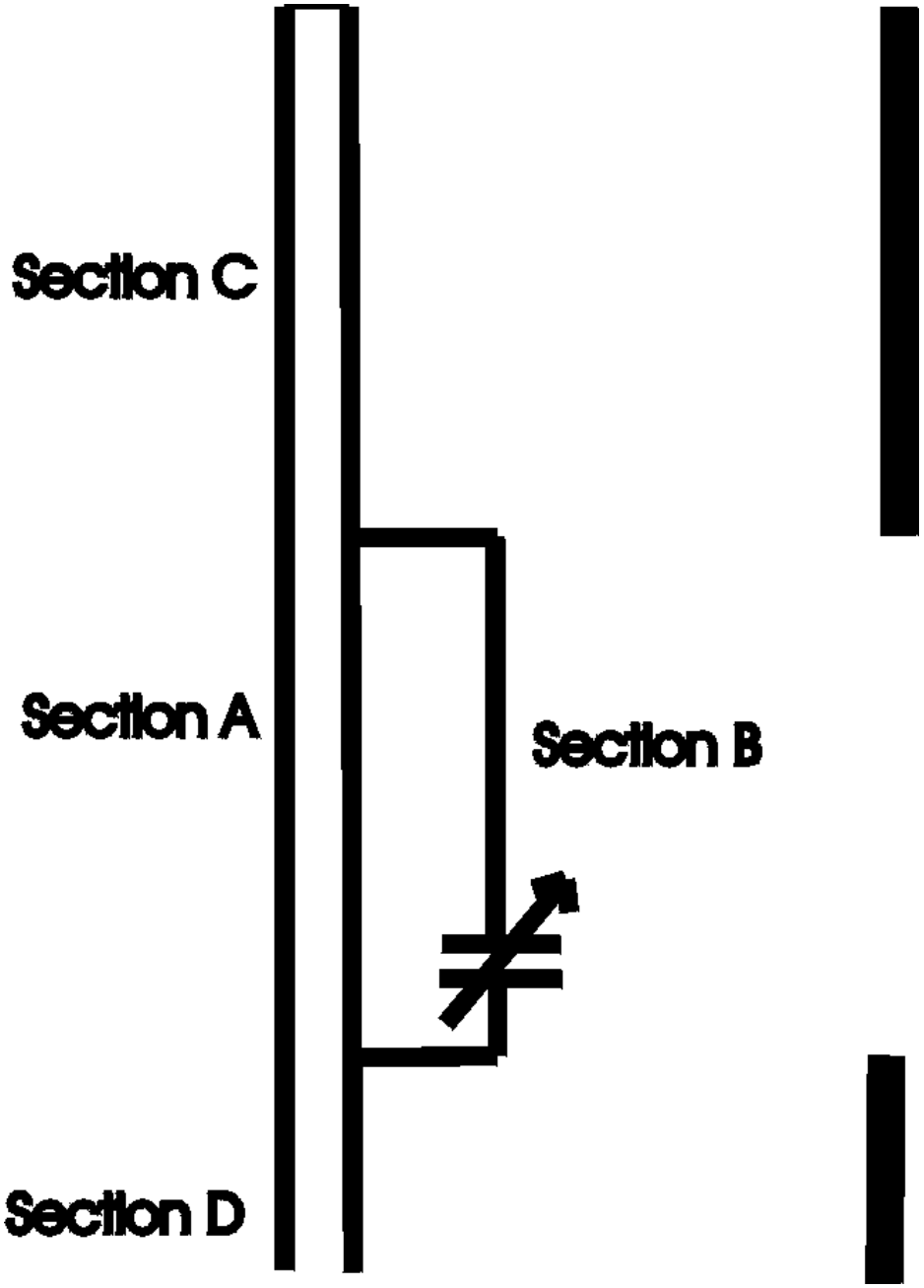
Detuning Towers

The goal of detuning is not to avoid resonance, but rather to minimize re-radiation and/or current in the interfering structure.

When large enough, structures can re-radiate and cause severe pattern distortion even when they are NOT resonant. Consider, for example, how effective a large non-resonant screen or solid disk is at reflecting signals even when it is nowhere near resonance.

We can minimize re-radiation by making an area or areas of the structure "electrically vanish". We often call this "de-tuning", even though it is more correctly electrical trapping or sectionalizing of a structure.

Most structures or towers, when detuned, have a section adjusted to represent a ***parallel tuned*** circuit.



Section A and B carry out-of-phase currents. Picture the current flowing upwards in A. It must then flow downwards in B. Since it is a closed loop, these out-of-phase currents are equal and flow in ***opposite directions*** at resonance. The result is radiation from sections A and B cancel each other. When section A and B are exactly resonant, sections D and C are isolated by a high impedance. The high impedance is caused by or related to the high current though the capacitor and the inductance of section A. When current is maximum, voltage drop is maximum.

This results in the electrical structure on the right, with section A and B removed!!

In effect, we have created a trap much like the trap in a dipole. As in the trap dipole, current is maximum in the trap at the trap's resonant frequency.

NOTE: ***The condition of proper tuning occurs with MAXIMUM current in section B, NOT minimum current!! To electrically sectionalize the tower and isolate C and D (and minimize radiation from A) section B must be tuned for MAXIMUM current!***

As either section C or D approach resonance by themselves, the tuning condition will change. This would occur when D is grounded and near 1/4wl or an odd multiple of 1/4wl long, or when C (with whatever is mounted on it) is self-resonant with section A removed!!

Under this condition, you would either need to sectionalize and detune C or D with additional detuning, or move the location of sections A and B to a new point that (when isolated) prevents resonance in C and/or D.

A few general rules apply. Pay attention to these guidelines to insure best results:

- Never parallel-tune a large area. Certainly not an area over 3/16th wavelength long
- The detuning "loop" must have a good solid connection to the structure being detuned. Don't connect the detuning wire out to a separate object or earth stake
- We want to adjust for MAXIMUM current in section B, the exception being when that would cause resonance in C or D
- We can not have any electrically large structures or wires hanging from the tower in the area being detuned.
- Ideally any cables passing the detuning area should be grounded to the tower at the top and bottom of the detuning area, or pass through that area in the center of the tower or mast. At the very least, cable shields should be bonded to the tower at the top and bottom of the cable run and unshielded cables placed inside the tower.
- Tuning is fairly narrow. ~5% total BW is about all that can be expected in most cases, but this varies greatly with the system including distances to the other affected antennas and the amount of pattern distortion tolerated

Note: I added the section about cables on Feb 17,2003. I'm surprised cables are often not grounded at the top and bottom of tall towers, and that unshielded control cables are not passed through the inside of towers. Cables should always be treated that way for lightning protection if for no other reason!

Capacitor Size

The amount of capacitance and the voltage rating of the capacitor is not easy to predict. The size depends on unwanted power levels that excite the detuned structure, the electrical characteristics of the detuned structure, and the Q of the detuning section. Capacitance values will be fairly high with short sections on lower bands like 160, almost certainly in the range of a few thousand pF for ~20 foot long sections. The exact value would depend heavily on dimensions of the A to B loop.

Voltages across the capacitor are generally not high, although they can be at times. The "loop Q" of A and B affects voltage, as does the amount of excitation and load presented by the impedances of C and D.

MFJ sells a clamp-on calibrated current meter that will not perturb the system. It is a cheap version of a current meter I designed. This is a calibrated meter with internal amplifier that measures current from a few mA to 3 amperes, not the uncalibrated RF-sniffer commonly sold. Some RF-sniffers, including those by MFJ, actually change the impedance and resonant frequency of the system because the pick-up transformers are not properly designed and terminated current transformers. Avoid loop-stick type current meters, since they measure ANY external field and can provide misleading results. Use a current meter that is directly inserted in line B, or clamps around line B with the closed core of a terminated current transformer. Use a meter that does NOT perturb the system when removed!

Lacking a current meter, it is possible to tune this system with a grid dip meter, by forming a small one or two turn coupling loop. As an alternative, the loop can can be broken at any point near the capacitor and a MFJ-259 or similar antenna analyzer connected in series. Proper adjustment is at the point were minimum impedance occurs. If that impedance is not low, you probably are not effectively detuning the structure.

Multiple Stacked Antennas or Tall Structures

When multiple stacked antennas are used, especially on a fairly tall tower, it may be necessary to sectionalize multiple points. Individual sections between antennas can be resonant, or appear electrically long.

If the tower or structure or any part of the structure or tower becomes resonant when section A is tuned to present a high impedance, then we need to move section A or tune it to some condition other than maximum current (resonance). Adjustments under this condition can only be made two ways:

- A sampling loop can be mounted on the structure 1/10th wl or more above or below section B and adjusted for minimum terminal voltage
- Field strength of the pattern can be plotted, and the structure tuned for minimum pattern distortion
- Never detune an area that contains large yagis or other electrically large objects, like long conductive guy lines, dipoles, or cables leaving the tower

Omega and Gama Matching

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Omega's and Gamma's

Impedance Limits

The best matching system for feeding a grounded Marconi element, unless we are very lucky or very careful in planning the installation, is a simple gamma capacitor and shunt feed conductor with the tap point adjusted to find 50 ohms resistive.

While losses are ***not*** significantly increased when using an Omega match with reasonably sized components, matching range is not significantly extended either! At best, the Omega match saves us a few trips up and down the tower while searching for the precise shorting strap position between the gamma wire or cage and the tower. If we are already at the top of a tower with the gamma tap point, and if at that point we find resistance (after reactance is cancelled by adjustment of the gamma) is too high, an Omega will not help. The Omega can only match loads ***LESS*** than 50 ohms resistive by stepping the resistance up! It can NOT step or transform antenna feed resistance downwards when using capacitors, and neither the Omega or Gamma can match capacitive antenna loads.

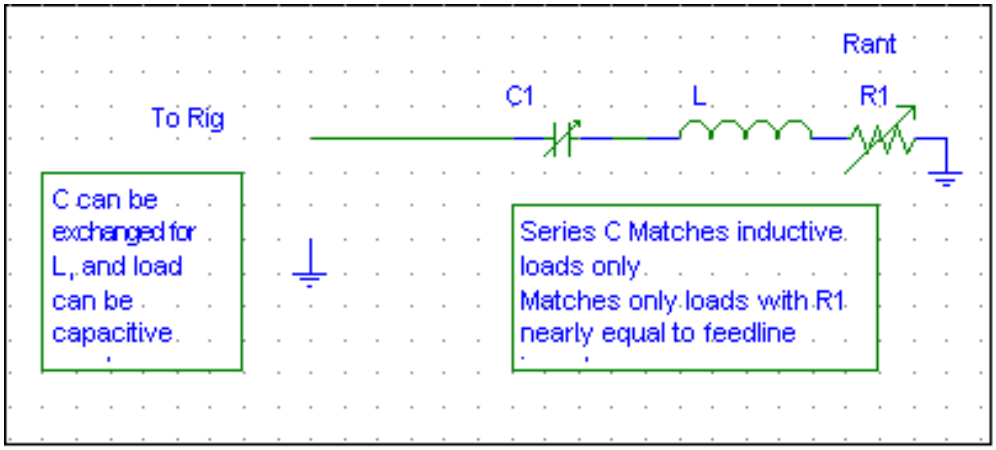
By adding an inductor, we can greatly extend the matching range. The extra time, expense, and complexity of adding a high-Q inductor would not offset the slight reduction in effort to simply find the correct shorting point for gamma and omega capacitor systems.

Gamma Match

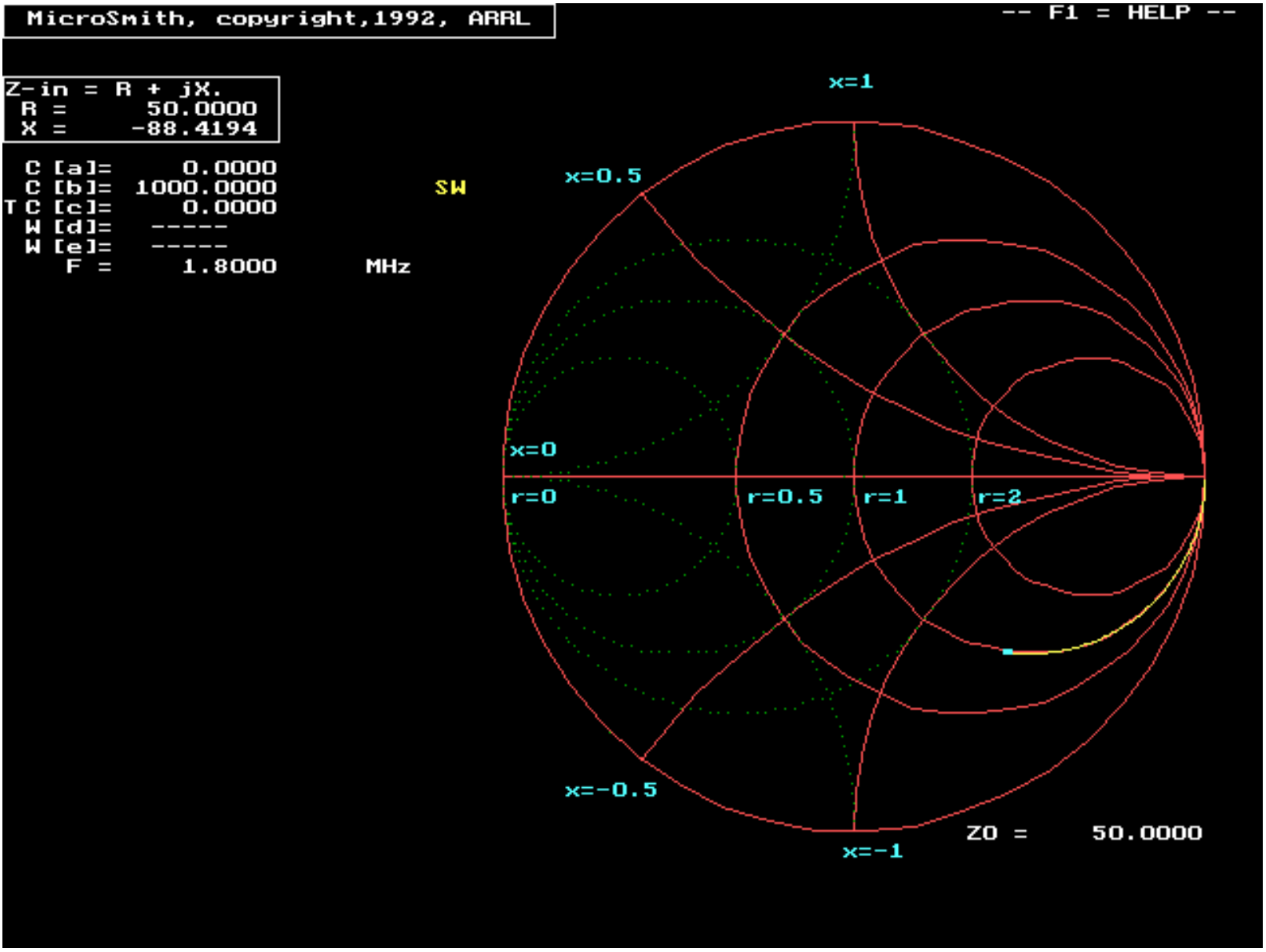
The gamma match capacitor can only cancel reactance, it can not modify the "real part" (resistance) presented to the feedline. It is the most simple form of matching, and has the lowest operating Q and loss of any system (when it is useable). Adjustment of resistance (real part) requires adjusting the diameter, spacing, or length of the gamma section.

The voltage rating of the capacitor is simple to find. It is Y*I*X where Y is the safety factor and adjustment for peak voltage (RMS*1.414), I is the RMS current from the feedline, and X is the reactance (in ohms) of the capacitor. Y is normally a factor of four to allow for peak conversion, SWR, component flaws, and transients.

The current rating is simply the RMS current plus a safety factor. Component physical size as well as Q control the current rating. See component ratings.



Looking at a Smith Chart, we see a single line (yellow) following a portion of the resistance circle with a resistance value of 1 (1 is normalized to 50 ohms). The gamma match can not deviate from that resistance circle, and so the resistance must be adjusted or transformed by other means. We normally adjust the gamma system's tap point by using a movable shorting strap to change R. Just moving a short is enough, it is not necessary to remove the extra "unused" portion above or beyond the shorting strap.



Let's consider a 1500-watt transmitter and 50-ohm transmission line, with a 1:1 VSWR. The matched line current is found using $I := \sqrt{\frac{P}{R}}$. The result is 5.5a RMS current.

With 1500 watts and 50 ohms we have 5.5 amperes RMS. Assuming the match requires 300pF at 1.8MHz we have: $Xc := \frac{1000000}{2 \cdot \pi \cdot F \cdot C}$ $Xc = 294.731$

4*295* 5.5 = 6490 volts. It's easy to see why arcing is a problem when capacitance is low! Very few air variables would take such a high voltage, and insulators (especially when exposed to moisture) also become a problem. This is typical of a shunt-fed element when the matching section is a thin long wire.

We can decrease reactance by increasing the effective diameter of gamma conductors, and this will increase required capacitance value. One way to decrease reactance is to use a cage of wires spaced as far apart as possible. ***Halving the reactance (doubling the capacitance value) will cut peak voltage in half, with no other changes. This has the effect of quadrupling the power rating when using the same breakdown voltage ratings in components!***

A thicker gamma conductor also lowers operating (loaded) Q. This is a series-resonant system, operating Q is set by the amount of reactance in series with the load resistance (50 ohms). Reducing operating Q has the effect of increasing bandwidth while simultaneously reducing loss and increasing power rating.

Note that bandwidth increased while efficiency increased. This happens in many cases. Popular folklore tells us narrow antennas are efficient, but that is true only in a few specific cases. In most cases, bandwidth by itself tells us nothing about system efficiency!

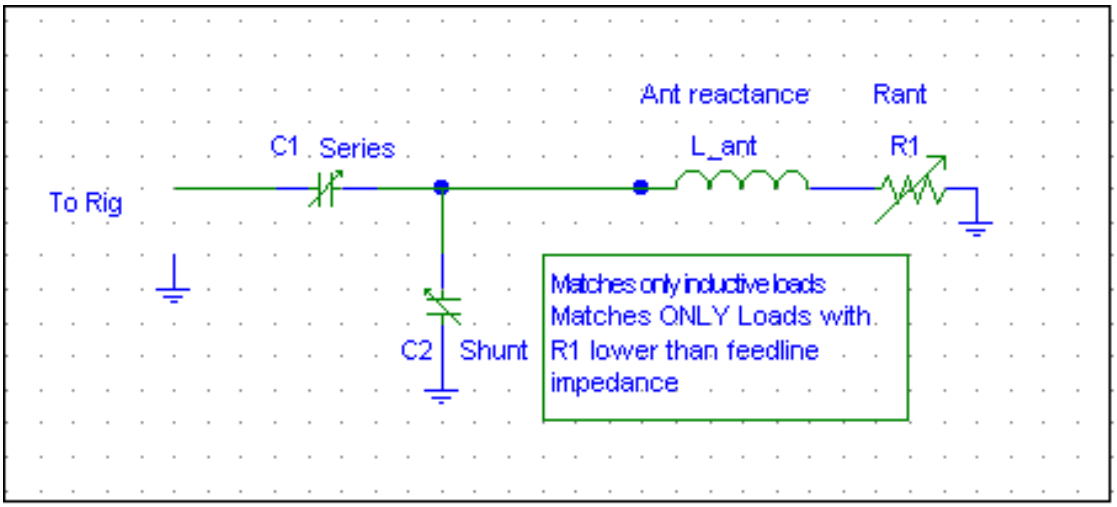
One example of a conundrum with gamma capacitors is found in yagi antennas. If we increase capacitor dielectric thickness by decreasing inner rod size, the voltage across the capacitor can actually increase at a faster rate than insulation thickness increases! The same is true if we increase gamma capacitor length in an effort to restore capacitance after increasing outer conductor diameter. The effect thicker insulation and higher voltage rating, if done incorrectly, can be to make little change or perhaps even ***decrease*** gamma capacitor power rating!! Always use a large diameter gamma rod, and increase the dielectric thickness ONLY by increasing the outer rod to accommodate dielectric thickness changes. At the same time, the inner rod should be increased to avoid any requirement of making the gamma capacitor longer. Avoid sharp points or edges on the rods, just as you would avoid sharp points in any HV system....especially inside the "capacitor".

The above example of decreased power rating is especially important to Amateurs using coaxial cables as capacitors. Voltage is NOT constant along the length of a long coaxial capacitor. Maximum voltage in the component is always ***HIGHER*** than the actual voltage across the terminals of the "capacitor", and it is higher than the voltage calculated by the current through the capacitor! Coaxial capacitors or linear stubs used as reactive elements always have significantly lower operating Q, higher power loss, and operate under more electrical stress than a well-designed lumped component. Stubs and linear loading does have the advantage of spreading heat out. You won't notice the heat as much, even though there is a lot more heat energy! Just don't let the smaller temperature rise fool you into thinking the system has less power loss.

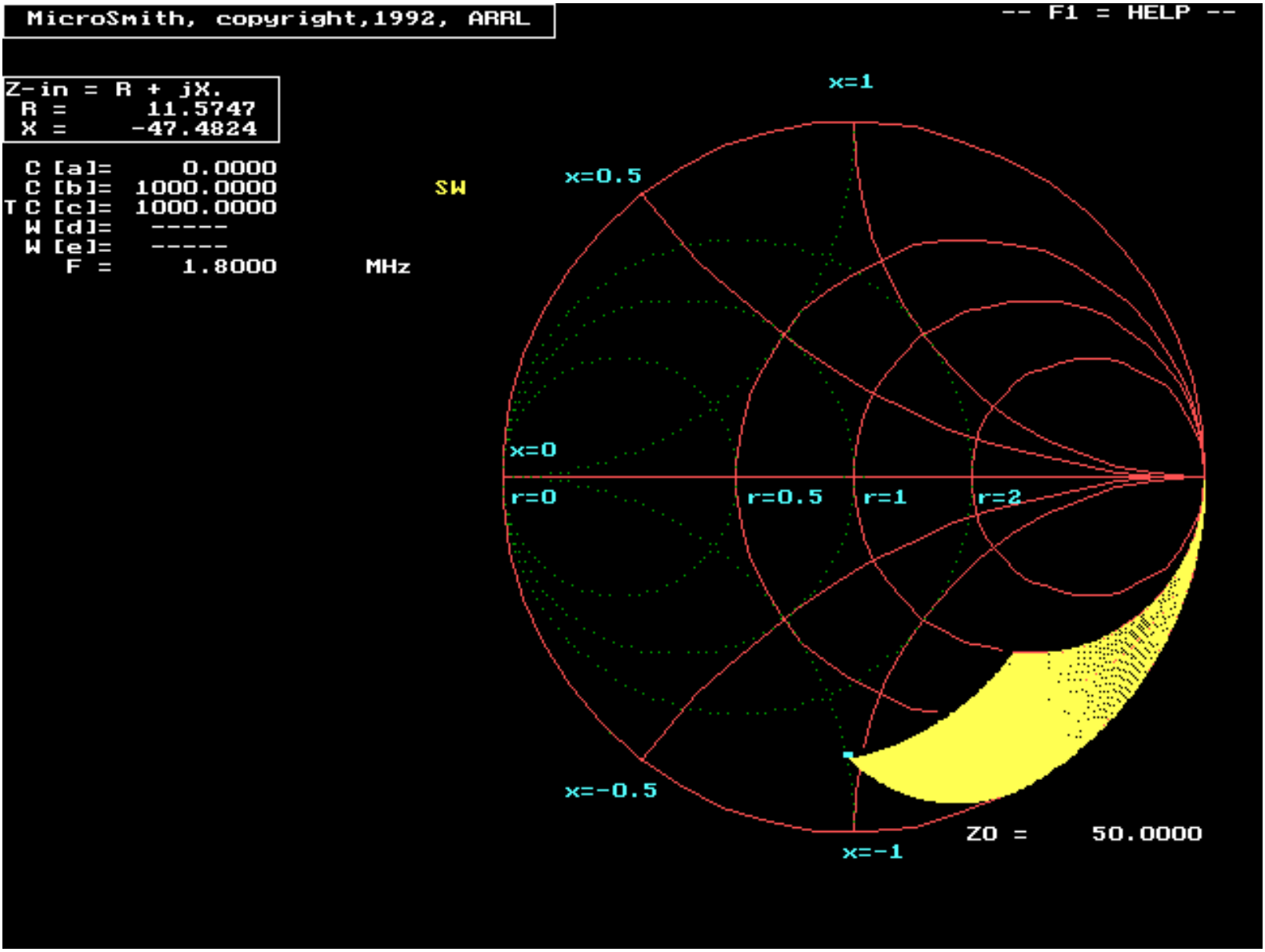
Omega Match

The Omega Match is really just a form of the simple L-network.

Traditional Cs/Cp Omega Match



In this system, C2 parallels the shunt terminal and primarily sets the resistance transformation C1 is in series with the feedline, and primarily cancels reactance. If we look at this network on a Smith Chart, we see the resistance range is somewhat expanded over a conventional Gamma Match.



We also can see it is not the panacea often claimed. Look at the "r" circles on the above chart. "r=1" is normalized to 50 ohms. With that in mind "r=2" is 100 ohms, while "r=0.5" becomes 25 ohms. The Omega's resistive matching range, using 1000pF vacuum capacitors, would only from 50 ohms ***downwards***. Voltage ratings are still a problem when the system requires capacitor C1 to have a small capacitance value (high reactance), and now we have two capacitors that have to handle essentially the same high voltages when C1's operating capacitance value is small!

This arrangement once again requires antenna shunt wire impedance to be ***lower than feedline impedance***. The network, since it only uses capacitors, ***must be used with inductive loads*** or the feedline terminal will be reactive.

One useful application of this circuit is in matching very short non-resonant vertical element, sometimes called a "Hairpin Monopole". Well-designed capacitors, such as vacuum capacitors, have such high Q they are essentially lossless. With vacuum capacitors and very good connections, losses in the matching system are very low even with very high currents. We shouldn't be misled into thinking we have a "free-lunch" magical short antenna! We "pay for our meal" in greatly increased conductor losses in the antenna conductor. The conductors not only carry common-mode radiation currents, they also must carry significant circulating currents involved with the feed-system. These high currents causes distributed conductor losses to be much higher than conventional

loading systems using a reasonable-Q lumped inductor. Bandwidth and radiation resistance for a given element size is often significantly less than conventional series-feed when using a hairpin monopole with Omega loading and matching, although the system has the advantage of being easily tuned to new frequencies. The Omega/hairpin vertical also allows the feedpoint resistance to be adjusted to nearly any value we might require.

C1 is primarily for power factor (reactance) canceling. It is rated exactly as in the gamma, voltage and current being determined by the required amount of operating reactance and the line current.

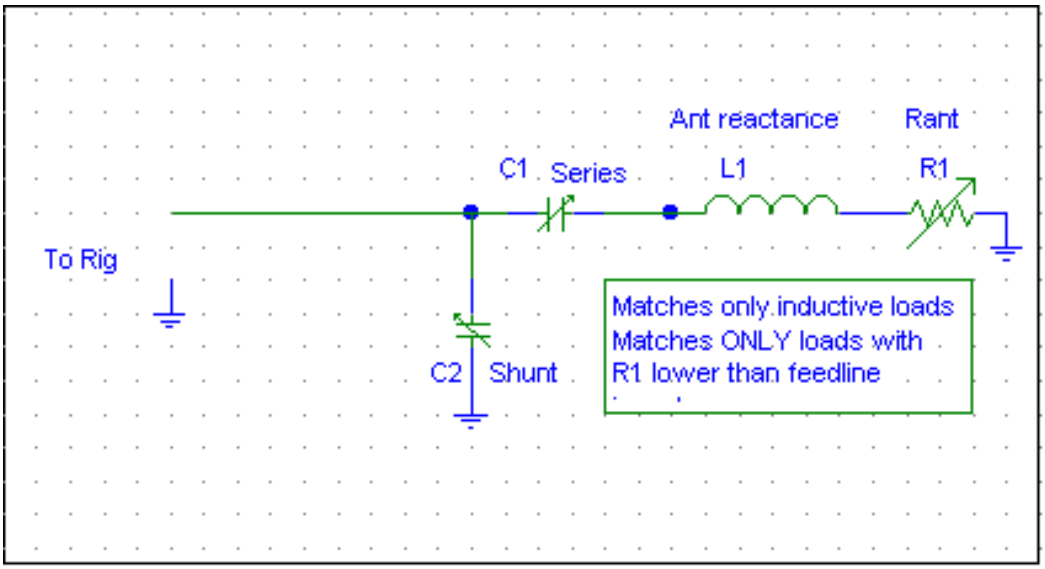
C2 primarily sets resistance transformation. The component's electrical requirements are found only by knowing the load impedance and the reactive voltage and current in that branch of the system. It handles a portion of the input current, as well as a portion of the shunt current, so the current in C2 is always higher than the current from the source flowing through C1. The required voltage rating of C2 is also always higher than C1.

Assuming we have 10-1000pF variables, the normal matching range covers inductive reactance ranging from 80 to 4000 ohms with a resistance between 1 and 50 ohms. Lower reactances are only able to be tuned when the resistance is near 50 ohms. As you can see, resistive matching range is wide but limited to values below 50 ohms.

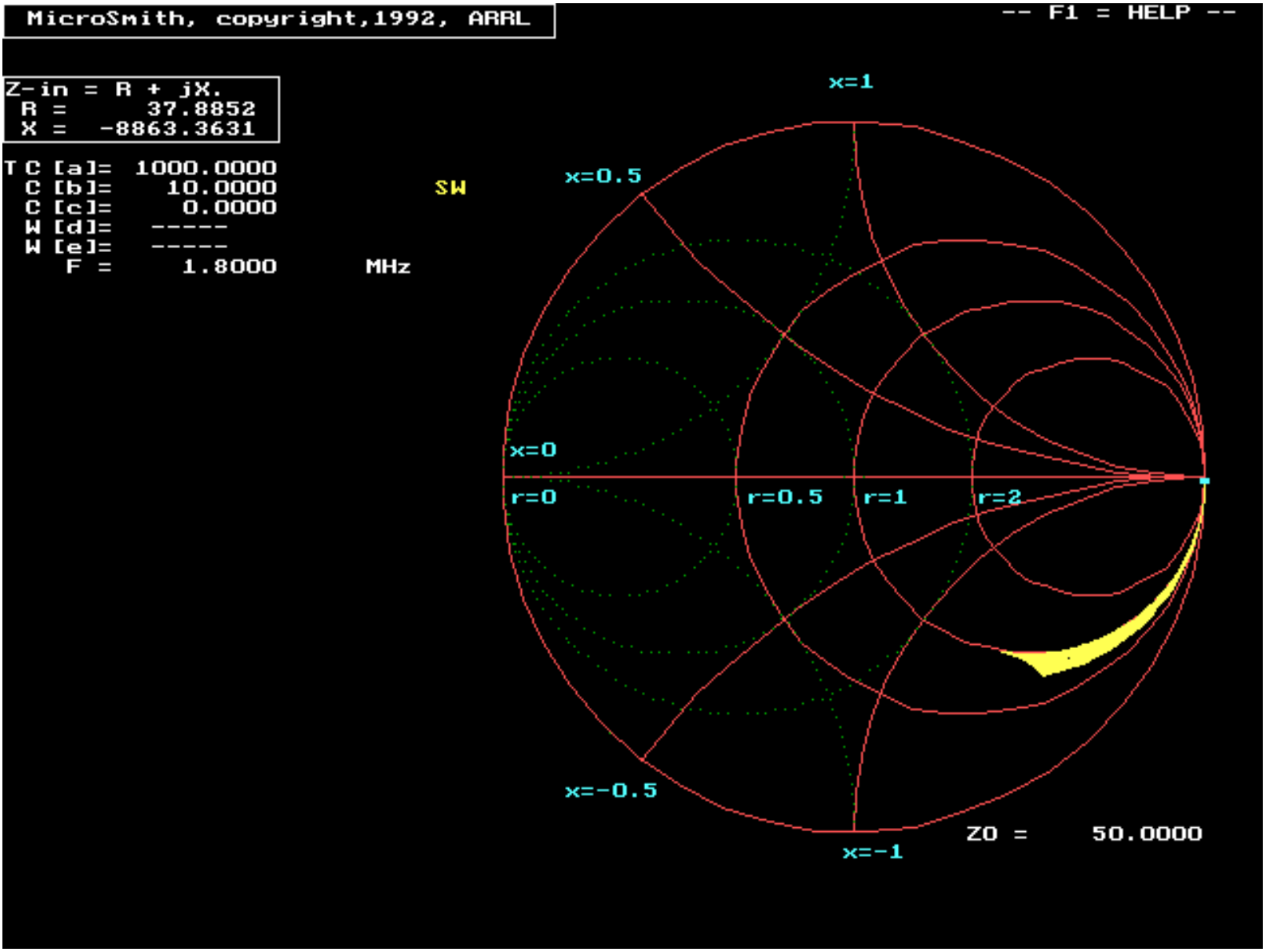
Cp/Cs Omega Match

The parallel-C input, series-C output is also a form of the L-network. The output, once again, must be the lower impedance while the feedline must have the higher impedance. Once again, this system only matches inductive loads.

It is not a very useful circuit, except in specific applications. The only advantage of this arrangement is C2 operates at significantly less voltage. The voltage across C2 is always no greater than the transmission line voltage, larger receiving-type air variables will operate at 1500-watt and higher power levels. Because C2 is under significantly less electrical stress, It is the better choice if you are very close to having a match and only need adjust resistance upwards slightly for perfect match.



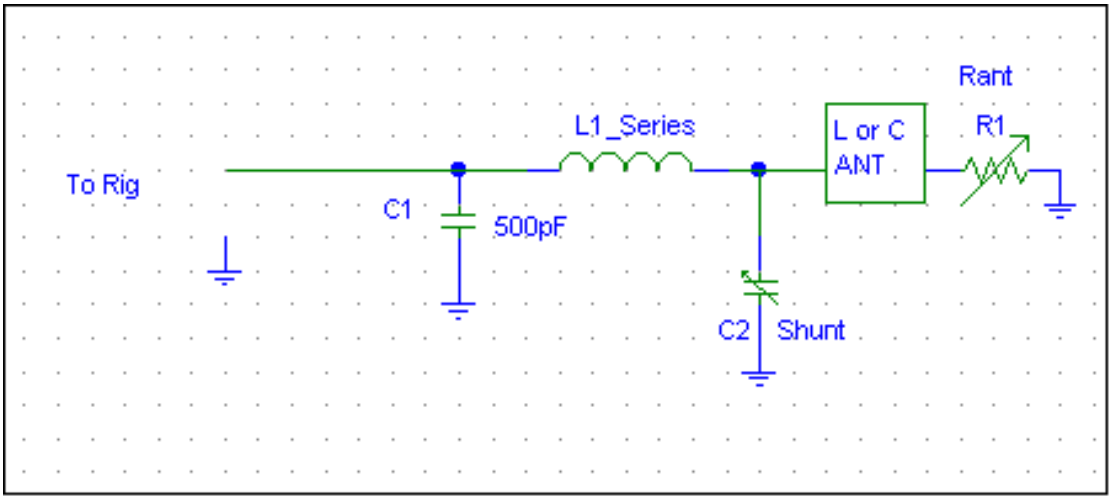
C1 primarily sets reactance, while C2 primarily sets resistance.



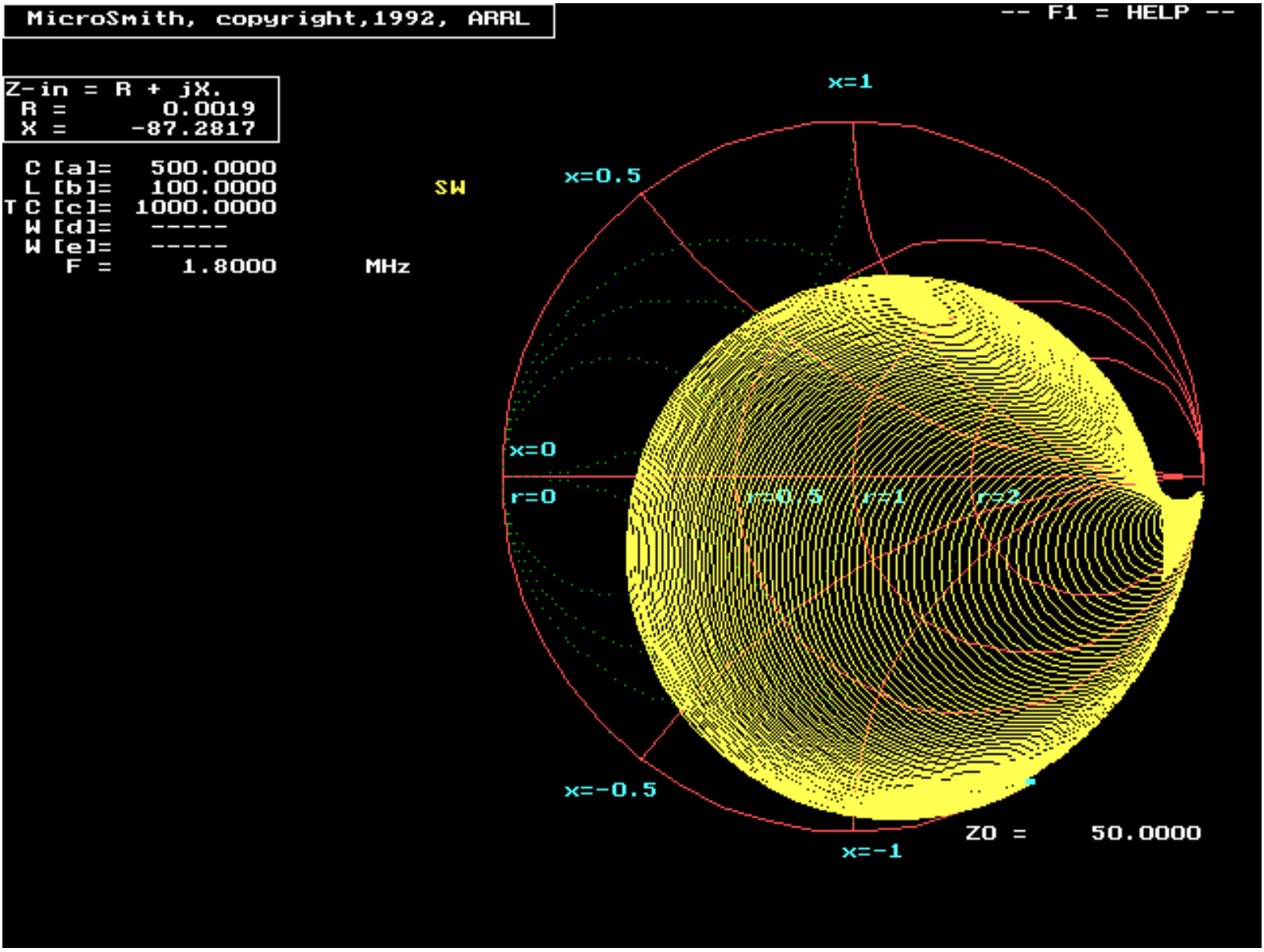
Assuming we have 10-1000pF variables, the normal matching range is reactances up 8000 ohms inductive and resistances between 38 and 50 ohms. As you can see, resistive matching range is very limited. C2 would need to be a very large capacitor (very low reactance) to have a wider resistance range.

Pi Matching

In cases where we can not find a 50-ohm tap point, we can add an inductor. In this case C1 can be a fixed or variable capacitor capable of handling modest voltage and current, such as a snubber-mica or receiving-type air variable.



L1 has to be fairly large to match high resistance and reactance ratios, at least 100uH on 160 meters for the extremes shown in the Smith Chart below. Assuming a 10-1000pF capacitor for C2, a 500pF capacitor for C1, and a 100uH adjustable inductor for L1 we would have the following matching range:



Matching range is greatly extended, and even covers capacitive loads.

Power Ratings

Heating is a long term problem that accumulates with time. Failure occurs when the heat causes a component to physically change from excessive heat. Heating is related to power dissipated by a simple formula $P := I^2 \cdot R$ The resistance is normally determined by knowing the component's Q (not the operating Q) and the current through the component.

The series resistance that dissipates energy as heat is found by dividing reactance by component Q. With 88 ohms and a Q of 5000 (typical of a good large air capacitor) Equivalent Series Resistance is 88/5000 or .0176 ohms. Heat is given by .0176 squared times 3.11 amperes. Heat is .0176*3.11*3.11 or .170 watts, not bad for an air variable capacitor. (Multi-layer chip capacitors can have Q's in the tens or hundreds of thousands, as can vacuum capacitors. Simple ceramic disc and small mica capacitors generally have Q's in the upper hundreds to many thousands.)

Current through C2 is found by dividing 274 volts by R-ant. If R-ant were 15 ohms the current would be 274/15 or 18.27 amperes. Voltage would be C2's reactance of 88 ohms times 18.26 amperes times the safety and peak factor of 4. Peak voltage would be 6500 volts for an 88-ohm reactance capacitor. This is a good example of why "T" antenna tuners arc with low load impedances while working fine at higher impedances, and handle more power when tuned to use maximum possible capacitance.

Arcing is an instantaneous voltage problem. In general a solid dielectric (insulation) punches through and fails almost instantly. Voltage rating is reduced with every arc, even brief unnoticeable arcs. As a general rule solid dielectrics suffer non-recoverable damage from any arc, even a very brief arc (such as a momentary low-current static discharge.

If the dielectric is a vacuum, liquid, or gas, and if the component's conductors or dielectric do not physically distort or change from heating, the insulation can "heal" and full performance is often restored. With an air capacitor, ionized air must circulate out of the capacitor.

If an arc causes a physical change, such as a hole, carbon path, sharpened or raised edge, breakdown voltage is almost always permanently reduced. One exception would be when an arc removes debris or sharp points, such as melting copper whiskers in a vacuum capacitor (copper can actually grow tiny whiskers in a high vacuum). Sharp points, even microscopic sharp points, greatly reduce breakdown voltage. If a sharp point is reduced and rounded by an arc, voltage breakdown will increase. If we actually melt the plate, voltage rating is reduced. Surfaces must be micro-polished, we cannot sand a capacitor plate to restore full breakdown levels if a capacitor is arc-damaged (variable capacitor plates are commonly polished by tumbling them in a very soft abrasive like walnut shells).

Capacitors in parallel with a known resistance. RMS sinewave voltage across the component is given by $E := \sqrt{P \cdot R}$ where P= maximum possible level of PEP applied. R=resistance the capacitor parallels.

Capacitors arc and fail from instantaneous voltage peaks, not from average or RMS voltage. We multiply the

RMS voltage across C1 times 1.414 and add an important safety factor. Multiplying RMS voltage by four normally works for outdoor mounted air-capacitors, as long as they are kept debris and moisture free.

A 50-ohm feedline with 1500 watts of absolute peak power (1500 watts CW or FM carrier or 1500 watts PEP SSB or AM) applied results in 274 volts RMS times 4 or about 1.1kV. Use a 1.1kV or higher voltage variable. This would be about .030" air gap in a typical construction air capacitor with normal manufacturing irregularities (generally very safe to assume .01" spacing for every 300 volts). With 50-ohm matched lines and a 1500-watt PEP (or CW carrier) transmitter, smaller transmitting or larger receiving air variables will work.

Heating

Current through resistance causes heating. The resulting heat over some period of time can cause physical changes that physically alter the component. In the short term, this can change the electrical parameters such as capacitance or voltage breakdown.

To find electrical energy converted to heat, we have to know the equivalent dissipative parallel resistance and voltage across the component or the equivalent series resistance (ESR) and current through the component. I generally work with current and ESR.

The current is a function of reactance and voltage across the capacitor, just like current through a resistance is found from voltage across the value of resistance. Assuming we have 1500 watts and 50 ohms resistive load directly across the capacitor, we have 274 RMS volts across the reactance of C1. This voltage, over the reactance in ohms, gives us the RMS (heating) current flowing through the capacitor.

Use the formula I = E/X. With a 1000pF capacitor (88 ohms on 160m) we have 274/88 or about 3.11 amperes.

Note: This is not exact because capacitors have complex series-impedances, but it is close enough for our questions as long as we have reasonably good capacitors. One common exception where this simplification will not work is with long coaxial stubs. Long coaxial stubs have a considerable amount of distributed series inductance. This causes a voltage increase as we move away from the feedpoint in an open stub, or a current increase as we move out on a shorted stub. This increase in voltage or current increases loss, decreases bandwidth, and increases effective loss resistance. Not only that, a stub (or linear loading) has higher distributed resistance because it has long conductors, rather than thick compact conductor area like a conventional component.

Loss and Component Failure

We often gauge system "loss" by the temperature of the component. We often assume a "cool" feeling component has low loss and a hot feeling component has very high loss. Using "touch or feel" temperature without considering size and ability to transfer heat is a good way to estimate component life, but it is not a good way to estimate system efficiency or component losses. Unless we carefully consider the complex issue of physical size and ability to transfer or dissipate heat to surrounding objects or air, we will have no idea how much power is actually lost in the component. Small components not only get hotter, they get hotter much faster than large components!

A physically long coaxial capacitor has a large surface area to dissipate heat. Even a very lossy coaxial capacitor

will have significantly less temperature rise than a small very low loss compact disk capacitor, because the capacitor concentrates all the heat into one very tiny area. We might form the impression we have an inefficient system because the small capacitor drifts in capacitance value or fails from heat, yet it is usually the cool-running stub that has much higher power loss!

This doesn't mean we shouldn't use a coaxial stub as a reactance if they are reliable, but it does mean we have to be careful assuming it is the best solution for loss.

Bandwidth

Bandwidth tells us nothing about efficiency, unless we understand the entire system in detail.

In general the largest bandwidth occurs with the fewest reactive components. Any unnecessary reactance will increase system operating Q. One exception is when multiple networks with low loaded (operating) Q are used to obtain a large phase shift, rather than a few components that would require high operating (loaded) Q.

Stubs act like a combination of many small series-inductances with multiple shunt-capacitors. Coaxial capacitors not only generally have more loss than reasonably-constructed lumped capacitors, they also have less bandwidth. A stub's reactance changes faster with frequency than reactance in a compact lumped component, because it contains "unnecessary" distributed reactances. The stub also has higher internal voltages and currents, because of the distributed reactances. A coaxial stub may be cheap and "feel cold" in operation, but we shouldn't delude ourselves into thinking it is less lossy than a small capacitor. Remember it isn't only power dissipated that determines temperature rise, we must consider the total area heat is distributed over. 1 watt of heat feels very hot in a component the size of a pencil eraser, while 100 watts of heat could be undetectable by touch in a long thick piece of coaxial cable!

Transmitting

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160 Meter Transmitting Antennas at W8JI

200-Foot 12-inch face Tower

160-meter vertically polarized transmitting antennas center around this 200-foot tower.

When I moved here in December 1998, this tower was hastily erected as a quick way to get on the air. A single 0.5" diameter 50-ohm feedline and 8-pair control line run to the tower. A large outdoor storage box at tower base holds 160-meter phasing and matching systems.

OMNI Pattern

This 200-foot tower is the main omni-directional antenna. It is series-fed against a ground system of about one-hundred 200-ft radials. The radials are shallow buried 16-gauge bare solid copper wire known as "buss wire". A small low-pass L-network using a vacuum capacitor and variable inductor matches the tower. A 50-foot transmission line coiled around an insulated form between the L-network and RCS-8V antenna relay detunes this tower when the four-square is used. The four-square is detuned when the omni-tower is used.

318-foot 18-inch face Tower

High and Low Dipoles are supported by this 318-foot tower.

Most of my dipoles are supported by this "dipole support tower". This tower is 318-foot tall and sits on a base insulator.

This tower started as an elevated radial test tower and grew from there. Because of the early elevated radial experiments, the 318-foot tower has the remains of 60 1/4-wl buried radials below it.

At the top: Our two-meter repeater antenna

Three nylon U-V resistant 7/16-inch ropes over pulleys for pulling up antennas.

Guy lines are EHS steel, with sectionalizing insulators. There are three anchor distances. Look at guyline slope.

At the 170-foot level, two more ropes over pulleys. Just below that, the middle marker lights.

Want to climb up?

View from 318-foot tower

We are only at around 250 feet here, and looking towards the east. The far right horizon has a water vapor cloud from a power plant near Juliette, Georgia about 15 miles away. Nearby is the Whistle Stop Cafe, from the movie Fried Green Tomatoes.

Ground height is about 800 feet above sea level, but the important thing for VHF and UHF is the ground slowly rolls away in three directions. The effective height above average terrain for your view for the next 20 miles is around 300-feet, even though you are only 250ft above ground.

The rope in this picture comes from the end of one of my 318-ft high apex Inverted Vee Dipoles for 160 meters, and goes out over those tiny 80 to 100-foot tall trees about 700 feet away.

My 28 Beverages and 12 vertical receiving antennas are out in the pastures, spread over a 1500 by 3000-foot area. They are controlled by BCD data over underground control cables. Eight coaxial trunk lines feed the receiving antennas by running to switching hubs, where smaller antenna feedlines leave.

For stereo recordings of 160-meter signals, click here.

Higher Bands

I'm embarrassed. Most of my work is on 160 meters, so I neglect the other bands. I'll work on 80-10 meters someday soon, I hope! At the moment, I only have a single rotating 160-foot tower that supports Yagis for 40 meters up through 6 meters. The hardware comes from K0XG, and has been trouble-free. It has sixty 140- foot long buried radials for lightning protection and to help with low NVIR antennas like dipoles. There are more antennas and feedlines on that tower now.

TX_four_square

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Four-Square

The 200-ft tower has five catenary lines, each about 500 feet long. Four of these lines, running NE, SE, SW, and NW support four 1/4-wavelength wire verticals used in a 160-meter four-square antenna.

Each four-square element is made from a "bow-tie" of #16 wire, which fans out 12 feet at the top. The elements are fed with 50-ohm RG-8 type coaxial cable. Each element has about fifty 1/4-wavelength radials. These radials are bonded to the large radial system of the 200-foot tower at points where more than two wires cross. This makes the entire grounding system for the 200-foot and four-square elements more similar to a large screen rather than just than individual radials.

Four-Square Design

One item overlooked throughout the history of the four-square is the effective element spacing. Effective element spacing is not the same as the distance along each side of the square. The effective spacing is less than that distance, because the array fires from corner to corner. The effective array length is 1.414 times the spacing along each side, not twice that length. With 90 degree per side array size, we really only have a 127 degree long end-fire array when firing across diagonal corners. The correct phasing for an optimum null at zero degrees elevation directly in line with the array is not 90 and 180 degrees, but is actually around 117 and 234 degrees! With the traditional 90 and 180 degree phasing, the array should be stretched to a corner-to-corner distance of 180 degrees, making each side 127 degrees long. Unfortunately feedlines of 1/4-wl electrical length won't reach that distance with velocity factor and connection lengths involved, so the obvious choice is to use greater than 90-degree phase shift. As an additional benefit, the directivity and gain are higher with closer spacing and larger phase shift. Performance is better for both receiving and transmitting.

The phase and spacing used are obviously not optimum, and the popular coupling units almost certainly do not do what we expect. They sacrifice gain because any four-square 1/4 wl on a side really does not "want" 90- and 180-degree phase shift, and they really are not designed for the real impedance presented by the feedlines from the elements. Why do we use such a design? Probably because it was originally used, and no one ever looked at the array in any detail. It was easy to copy the original concepts, and no one thought to check and see if the phasing systems were optimized.

Array Transmission Lines

(the following assumes lines have negligible loss)

I use 1/4-wavelength 50-ohm lines feeding each four-square element. 1/4-wavelength lines have a unique property. When ***odd-quarter-wave*** lines are sourced from ***equal voltages***, the load ends of the line ***will source equal currents into any load impedance.***

There is a second important often overlooked property. When a line is misterminated and has standing waves, total phase shift along the line does not equal the electrical length of the line. An exception occurs when the line is any multiple of 90-degrees.

Just as impedances invert every 1/4 wl, the fixed source voltage at one end becomes a fixed source current at the other end. Not only that, phase shift is constant even though the lines are terminated by different load impedances. Many arrays use this concept to advantage, and a current sourced 1/2 wl line would work just as well as the 1/4 wl lines if we fed the line with equal currents instead of equal voltages. (As a matter of fact that concept is behind a two-band transmitting array I designed.)

Current Distribution

A two-element directional system requires *equal current* at the current maximum of each directional element for maximum F/B ratio and gain. A three-element array would require a binomial 1:2:1 ratio for best pattern. The center element, in effect, has to supply a field for each of the two outer elements to work against, it is doing double duty. The four-square is a novel approach to a three-element array, the center two elements are fed in phase, and act as one common element. If all four elements have equal current, total center element current is twice the current in each end. The 1:2:1 current distribution is satisfied.

With elements series-fed at the point of current maximums, each element requires equal current at the feedpoint. (If the elements were voltage-fed at the voltage maximum, they would require equal voltages at the feedpoint to have equal currents in the elements!) This makes transmission-line type phasing systems easy to design. We simply design the system so each phasing-line has unity standing wave ratio, and tie the elements directly to the phasing lines. Because voltages are equal everywhere in the phasing system, element currents are equal at the end of each 1/4-wavelength feedline.

Phasing Systems

My design operates the phasing line without standing waves. Since the phasing line is not a multiple of 1/4-wl, if the delay line were misterminated phase shift would not equal the line electrical length.

Voltage at each end of the delay line would also be different. Since the line is matched, phase shift will be equal to the electrical length of the line and voltages will be equal at each end of the delay line.

There really isn't any need for "dump" resistances or hybrids in any phasing system. If you look at the systems, the dump resistances really don't guarantee performance. A four-square system never requires controlled "power splitting". As a matter of fact, equal power in each element would guarantee poor F/B ratio and gain.

Keep in mind that each element, other than the two in parallel, has a different feed impedance. With different load impedance at each element current can't be equal in each feedline element group if power is divided evenly. The four-square system, at least one with 1/4 wl feedlines, actually requires equal voltages. Equal power insures a poor pattern, while equal current at the element (or equal voltage at the start of the 1/4 wl feedlines) guarantees optimum patterns. Most four-squares work as well as they do only because the array is somewhat forgiving. They are a good example of reasonable performance through oversight, rather than careful planning.

I solved these problems in my system by placing the phasing common-point in the direction of maximum radiation and using crossfire phasing. This is very similar to the feed-system of a log periodic, except I optimized phase shift and current distribution.

A crossfire delay line system feeds the center two elements in my array. This is similar to the scheme I use in my phased receiving antennas. Each center element has around 25 ohms of operating impedance, and that produces 100 ohms at the end of the 1/4-wavelength 50-ohm feedline that goes to each element. The center two elements, when paralleled, have a combined impedance of 50 ohms.

The delay line from the common point to the center two element junction is 60-degrees of RG-8 coax. The 180-degree "flip" comes by using a choke balun. This is similar to the technique I use in receiving antennas. The primary difference is in core size and materials, my high-power inverting transformer uses eight 2-inch diameter 65-material cores and Teflon insulated coaxial cable. This choke allows the shield and center to be transposed at one end of that delay line. Loss is low in the delay system, because it is a matched 50-ohm line. It is also only 60-degrees long and only handles only about 2/3 of total system power (almost 1/3 of the system power is applied directly to the front element).

The rear element offers a slight problem, because that element has a very low impedance. It almost has enough current, through mutual coupling alone, to be fully functional. As a consequence, it requires very little common point power. With a 1/4-wavelength feedline and parallel reactance canceling, the rear element presents several hundred ohms of resistance to the phasing network. My solution for feeding the rear element was to build an artificial transmission line using lumped components. For the rear delay line, I used multiple L/C sections to simulate a transmission line

with a characteristic impedance matching the rear element. This has three advantages:

1.) Q is low, making phase-shift much less frequency critical.

2.) Voltage is equal at each end of the artificial transmission line, allowing the delay line to all parallel each other at the conmmon point.

3.) I can easily "tweak" delay line characteristics with a few adjustments and optimize the array null.

In effect, I created a high-impedance transmission line from individual L/C components, and only needed to use 120 degrees of lag (the 240 degree lead I needed is electrically the same as a 120 degree lag) in the network. This is all very similar to the system in my small receiving four-square. Because current is low, components can be modest sized. The end result is more bandwidth, more stability, and less loss than a simple one-stage network.

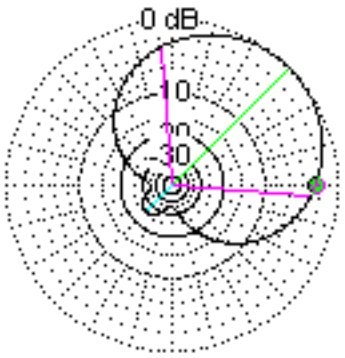
The Results

While this may sound like a complex system, it is very simple in practice. This array has exceptional bandwidth, exhibiting useful F/B ratio and gain over the entire 160-meter band. This also means a similar array for 80-meters would cover almost all of 80 and 75 meters. Additionally, this array has very low loss regardless of operating frequency. It does not divert power into a dump resistance.

My phasing design produces higher gain than conventional 90/180 degree phased four-squares through a tighter pattern and less loss.

Becuase the pattern is "tight", I added a simple system that provides a wide East, South, West, or North pattern. That pattern nearly equals the gain from a traditional four-square, and provides equal or better gain than traditional four-squares in all directions over a wider frequency range.

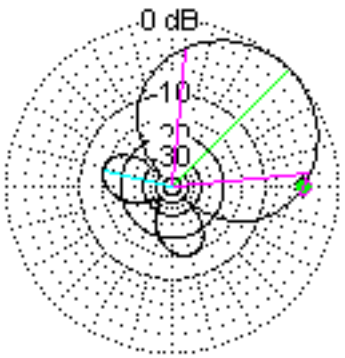
Here is the pattern of a traditional four-square at 15 degrees elevation, assuming perfect distribution and phase:



Azimuth Plot		Cursor Az	0.0 deg.
Elevation Angle	15.0 deg.	Gain	5.52 dBi
Outer Ring	8.0dBi		-2.48 dBmax
Slice Max Gain	8.0 dBi @ Az Angle = 45.0 deg.		
Front/Back	25.61		
Beamwidth	98.4 deg.; -3dB @ 355.8, 94.2 deg.		
Sidelobe Gain	-17.61 dBi @ Az Angle = 225.0 deg.		
Front/Sidelobe	25.61 dB		

Note the gain is 8dBi, and the 90-degree azimuth gain is -2.5dB maximum or a net gain of 5.5dBi in the "between" directions

Here is the pattern of the same array with 120/240 degree phasing:



Azimuth Plot		Cursor Az	0.0 deg.
Elevation Angle	15.0 deg.	Gain	5.13 dBi
Outer Ring	9.14dBi		-4.01 dBmax
Slice Max Gain	9.14 dBi @ Az Angle = 45.0 deg.		
Front/Back	37.28		
Beamwidth	78.8 deg.; -3dB @ 5.6, 84.4 deg.		
Sidelobe Gain	-5.48 dBi @ Az Angle = 166.0 deg.		
Front/Sidelobe	14.62 dB		

The gain is 9.14dBi, and the cursor azimuth gain is -4 dB maximum for an "in between" gain of

5.13dB. You only give up .4dB FS at the worse point for a net gain of 1.14dB in primary directions.

Gain increase through the phasing change is just over 1 dB through pattern changes, and is more than that through decreased loss (the model neglect phasing system and feed system losses. F/B ratio change is quite noticeable, as is the ability of the antenna to receive.

My actual working array is even more flexible, allowing the choice of eight directions plus omni.

In concert with a large ground system (the array sits in an area that has ground systems extending for at least a wavelength in all directions) and reduced loss in the phasing system, overall performance has been enough to offset salt-water path and distance advantages of northeast stations using smaller systems. Better F/B while achieving higher gain and more bandwidth is a definite advantage, although any one of the three might have caused me to use this system (besides the satisfaction of making something work as well as possible).

Baluns on Log Periodic Antennas

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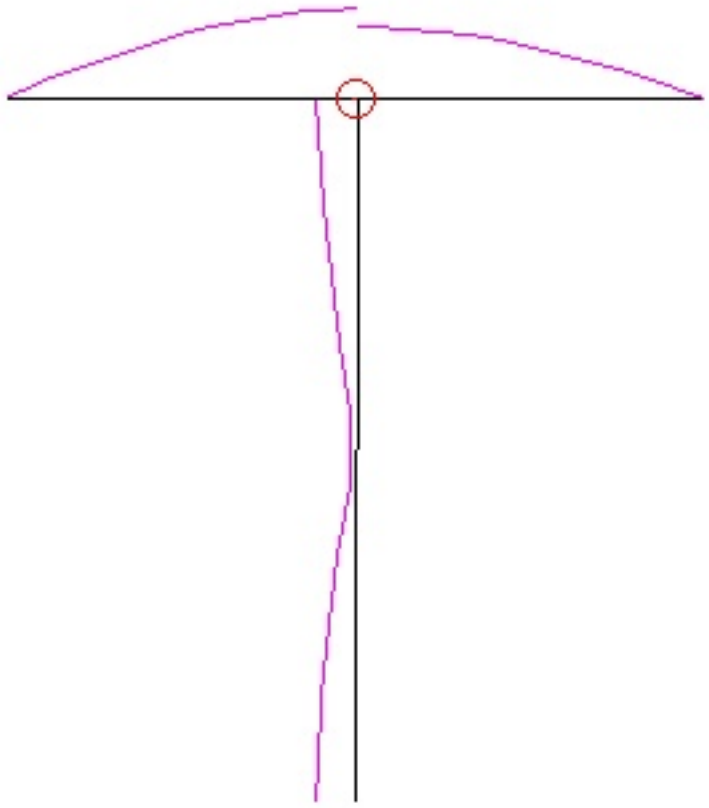
Some antenna manufacturers place baluns at the incorrect location in LPDA arrays. If we consider what the balun does we can see how the mistake occurs.

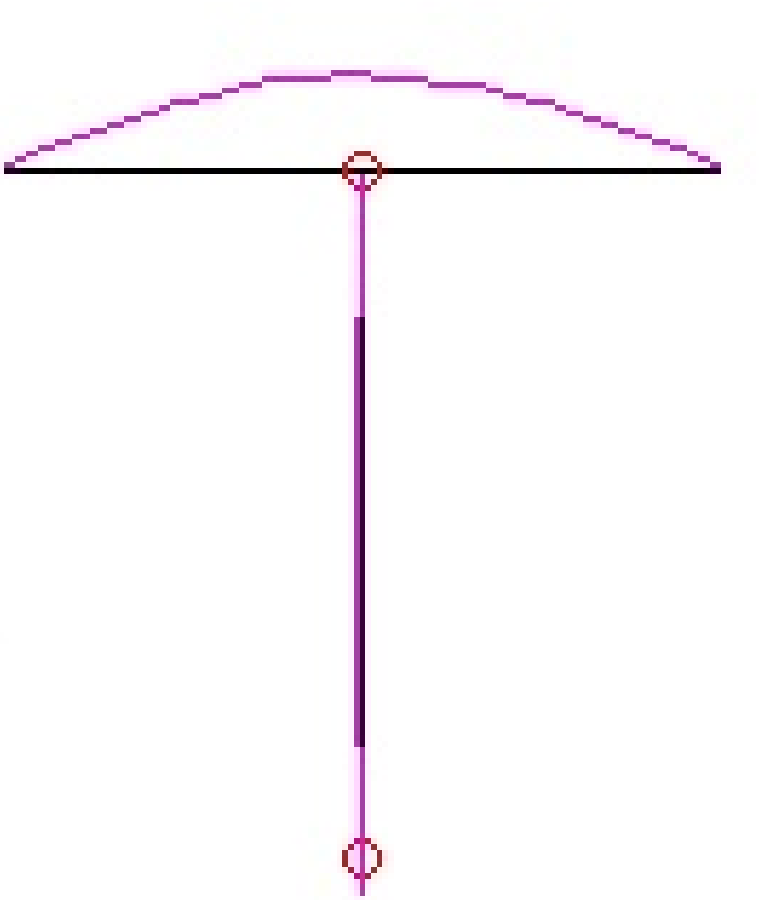
In the case of antennas, a balun isolates the shield of the cable from excitation by voltage differences between the feedline's termination and ground. In one case the termination might be a balanced dipole, with each terminal connection terminal having voltage to "earth". If the dipole is balanced and has 100 volts across the feed terminals, the feedpoint would have about 50 volts to an imaginary ground at the antenna.



The above current is for a 1/2 wave dipole 1/2 wl above ground, with coaxial feed. You can see the

feedline shield has significant current. In this case about 40% of the antenna's maximum current!

We can simulate a choke balun by adding a current source in series s with the shield, and setting current for zero amperes. The voltage across that current source will indicate the common mode voltage exciting the feedline.



Adding a perfect choke balun, current on the feedline shield goes to zero and the voltage across the choke balun is now found from the source menu:

Source 1 Voltage = 61.02 V. at -0.01 deg.
Current = 1 A. at 0.0 deg.
Impedance = 61.02 - J 0.009274 ohms
Power = 61.02 watts

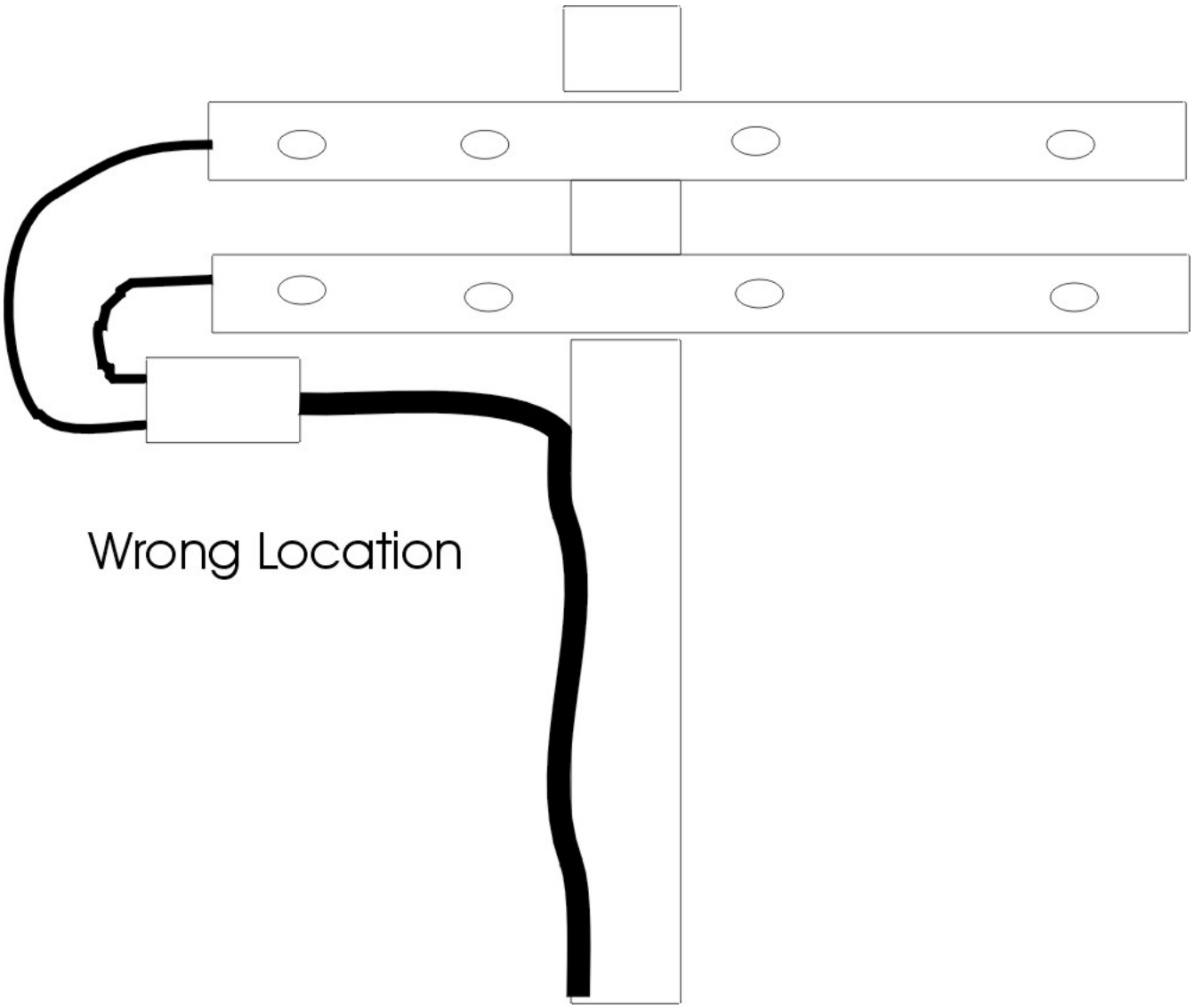
SWR (50 ohm system) = 1.220 (450 ohm system) = 7.375

Source 2 Voltage = 37.42 V. at 173.27 deg.
Current = 0 A. at 0.0 deg.
Impedance is infinite
Power = 0 watts
SWR (50 ohm system) > 100 (450 ohm system) > 100

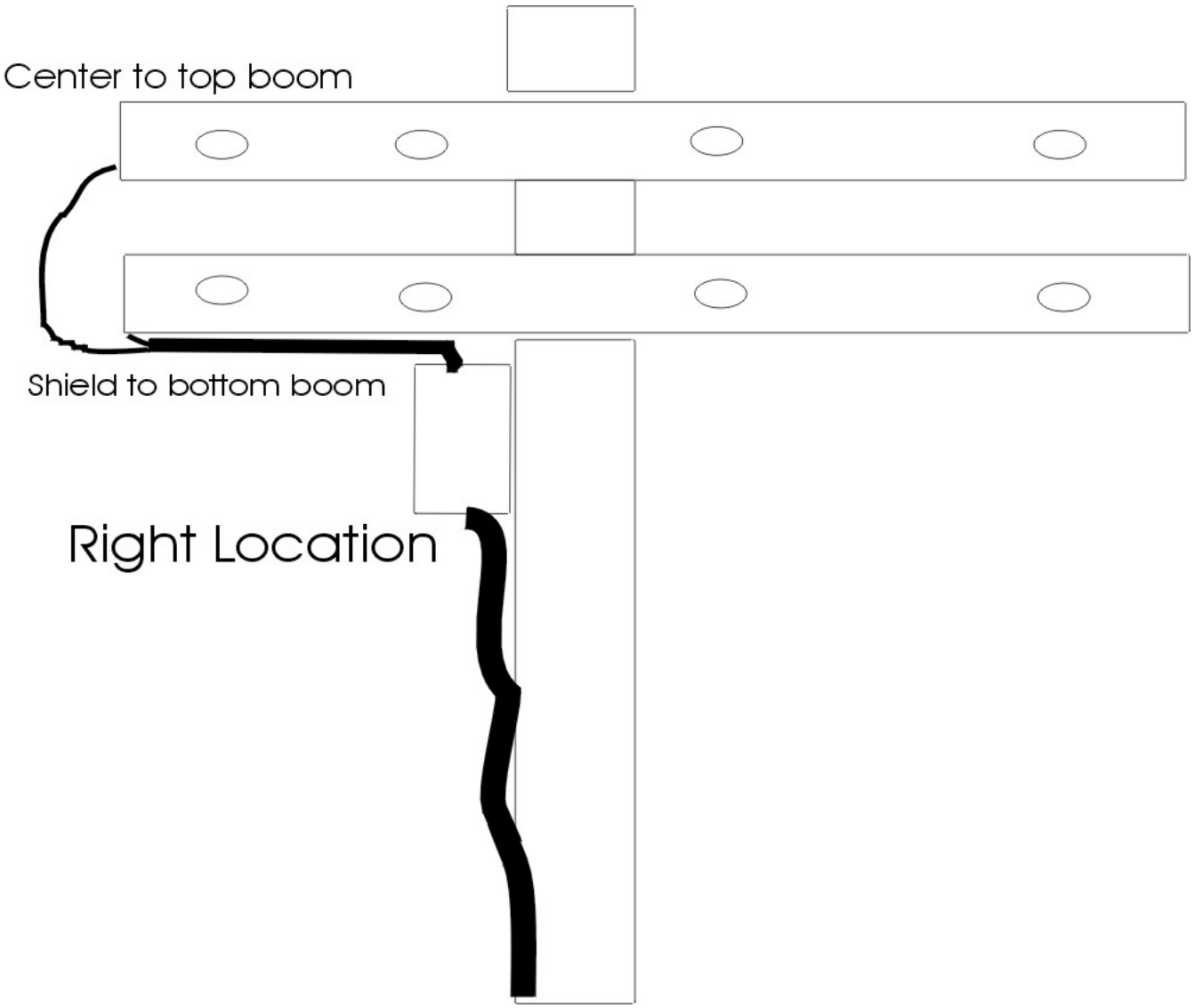
Source 1 is the actual terminal excitation of the dipole, while source 2 is the source that cancels common-mode feedline current. We can see the voltage across the perfect balun is indicated by source 2 as 37 volts, slightly more than 1/2 the feedpoint differential voltage. (We could also use a very high impedance load here, and do the same thing.)

What's Wrong with the LPDA Feed?

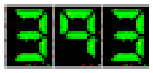
Some antenna manufacturers have you tape or otherwise attach the coaxial feedline along the length of a boom that is electrically hot, the boom feeding half the elements!! The very fact the boom feeds elements of the antenna requires the boom to have significant VOLTAGE along its length. Any cable parallel to the boom is excited by boom currents and voltages. Cables routed against the hot boom will have whatever the voltage is between the boom and the tower trying to excite current in the feedline shield.



If the log has a 1:1 balun at the feedpoint, you can simply tape the coax to the boom that the shield connects to without any balun at the log feedpoint. The best thing to do is to ground the coax shield to the hot boom at the exit point where the coax leaves the boom, and install a balun at that point! You can do this by installing a barrel feedthrough connector at the exit point and clamping the connector to the boom, then waterproofing the assembly. The balun goes between that connector and the feedline exiting the antenna boom and going to the station.



If the beam has an impedance matching balun at the feedpoint, or if you simply want to leave the balun at the feedpoint, you can leave the balun at the feedpoint. The feedline has to be suspended a few feet below the hot booms.

 **since 6/21/04**

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Snap on core material data

[Home] [Up]

Can anyone help me identify this core?

It is just over 1" long and about 1/2 inch diameter and snaps easily over RG58 or RG59 cables.

Please reply to W8JI at contesting dot com thanks!

Short dipoles and problems

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Recently I've been fielding problems from people using what has come to be called a "Cebik dipole". I also see people recommending this antenna to people with limited space for antennas. The problems I hear about generally center around tuner damage or failure of a tuner to match the antenna. An analysis of various installations has shown that the Cebik dipole is too short to be an effective antenna on the lowest band without some form of low-loss reactance compensation in, at, or near the antenna.

While pattern looks excellent on higher bands, lowest band performance of a short dipole (in this case 88 feet long on 80 meters or 44 feet long on 40) certainly leaves a great deal to be desired when feedline and tuner losses are included. It isn't the efficiency of the antenna that is in question, it is feedline efficiency and high voltages in the matching system that become problems.

The text below actually applies to any short dipole. If you don't like the dB or so loss of a G5RV system (including tuner and feedline losses), you ***really*** won't like an antenna shorter than a G5RV.

Loss in a Short Dipole

Power loss in short dipoles primarily comes from compensating reactances and matching to 50 ohms. This is because VAR (volt-amperes reactive) power is very high.

In a system with reactance, current and voltage are not in phase or in step. Because maximum current does not occur at the same time as maximum voltage, the simple product of current times voltage (I*E=P) can be much higher than the actual useful power. The higher reactive current causes increased current squared times resistance (I^2*R=P) heating and loss. The higher reactive voltage, in a similar fashion, causes increased dielectric losses.

In the worse case power loss caused by increased voltage and current in reactive systems can actually cause component failure.

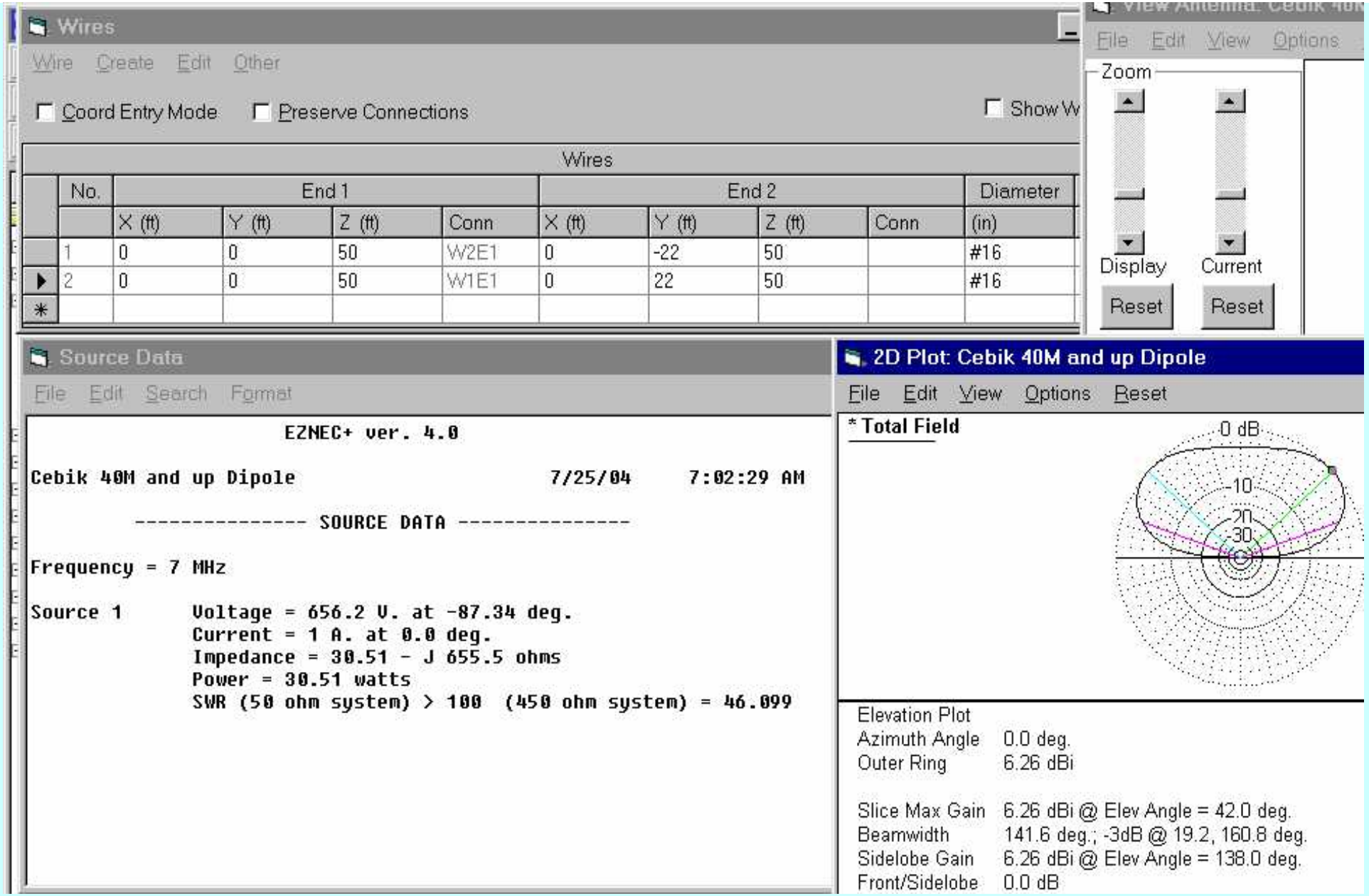
Short dipoles and verticals will almost always model with very good efficiency, because a model is a perfect lossless system if we do not include feedline losses and matching losses.

Let's look at an example by looking at the suggestion of shorting a dipole to approximately .33 wavelengths.

Dipole Impedance

Eznec shows a 44-foot dipole to have an input impedance of

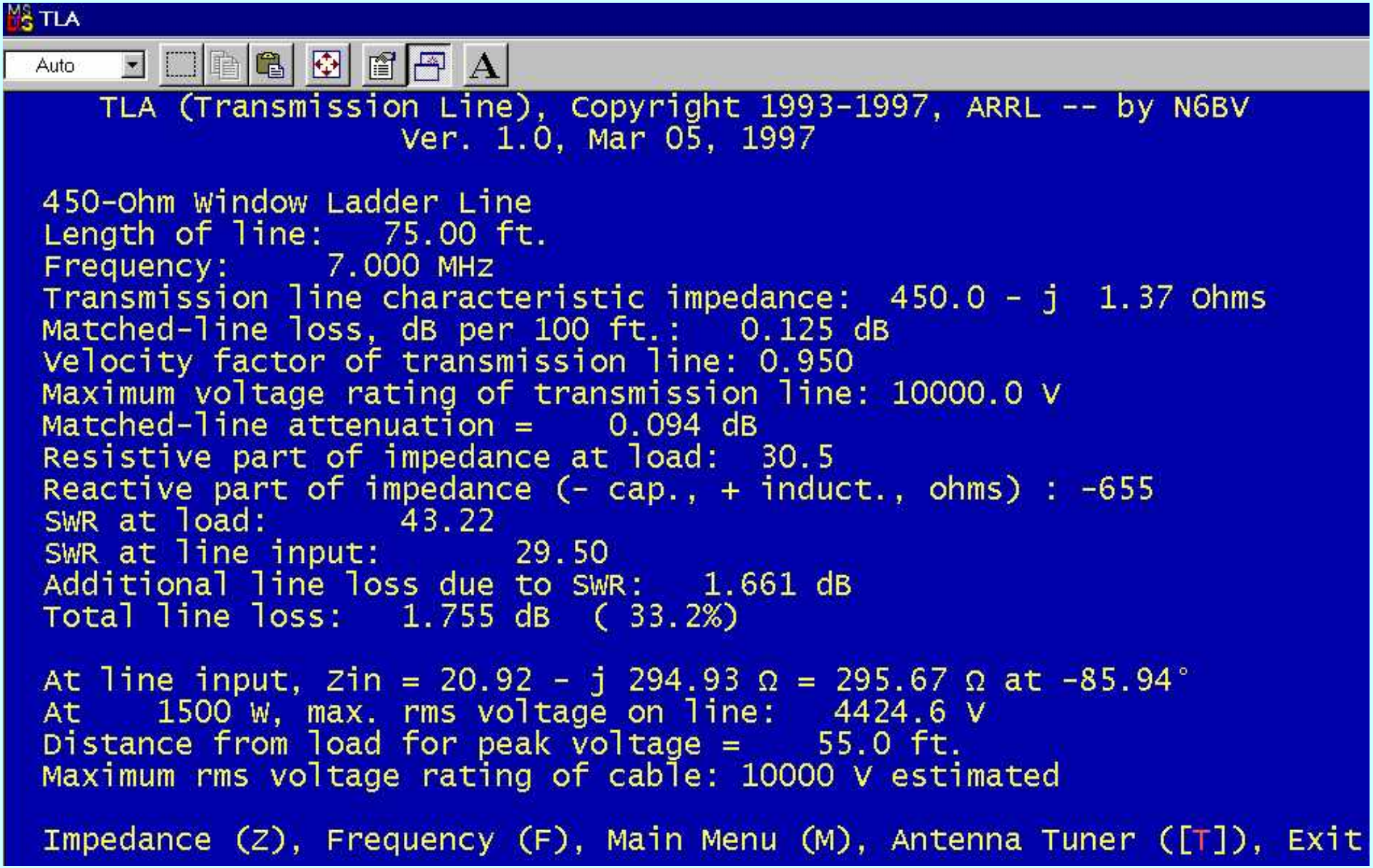
The model is shown here:



From this we see the impedance is 30 ohms resistive, not too bad. The problem is the reactance of -j655. This results in an SWR of over 47:1 on a typical "window ladder line", or 46:1 on a real open wire 450 ohm line.

I've verified the results of TLA (supplied with ARRL Antenna Handbooks) for dry new transmission lines, and it's pretty close. Unfortunately 450 ohm window line is a little worse than TLA predicts, and the impedance is generally lower. But we can consider TLA to be close.

Using TLA we find the following feedline losses for a 75 foot long feedline:



Note the following from the above:

- Line loss is now just under 2dB for a perfectly dry new feedline (it's actually just over 2dB if you use real losses, rather than TLA). This does not include tuner losses or less than perfect installation.

- The maximum RMS voltage on the transmission line is a staggering 4424 volts with 1500 watts. This is over 6250 volts peak voltage. (Even with 100 watts peak voltage would be 1600 volts!)
- Line input impedance is 21 -j295, this is what the tuner must match.

Suppose we add a typical "3KW" tuner to this system. Typical tuners have the following component values and Q's:

- Capacitors: 4.5kV, 250 pF, Q=5000 at mid-setting
- Inductors: maximum inductance 38uH with Q=200

We will assume a perfect balun without any loss. (It makes very little difference by the way if the balun is on the tuner input. Don't think moving a current or choke balun to the input of a tuner is a solution. The balun is under the same common mode stresses.)

With 250pF maximum capacitance, matching the odd impedance of this short dipole requires a tuner with the above component limitations to operate with a Q of 30.4. The resulting loss is 233.6 watts, and maximum voltage across a capacitor is 4250 volts! You are almost at the voltage rating of 4.5kV, so any small imperfections will cause a capacitor to arc. The inductor in the tuner would also dissipate 192 watts, which would damage most inductors unless duty-cycle was very low.

Some tuners (like MFJ Tuners) would be worse than the above values, Dentron tuners would be close to the above values, and many tuners would not even be able to match the load impedances. Automatic tuners and pi-network tuners would have a difficult time matching an impedance like this.

Conclusion

The lower limit in size of a multiband dipole, before feed system and matching losses move to the edge of severe problems, is about 200 feet on 160 meters, 100 feet on 80 meters, 50 feet on 40 meters, and so on.

A good rule of thumb is to multiply the band in meters by 1.25. The result is the minimum dipole length you can use without using a good matching system in, at, or near the antenna!

The G5RV length of feedline and antenna is the lower limit in size. A normal G5RV system, including tuner, typically has about 1dB of loss on 80 meters and less than 2dB ***system*** loss (*including loss from coax and matching*) on 80,40, and 20 meters. People seem to hate G5RV's, yet they now seem to be willing to further shorten the G5RV and recommend others do the same!

As an antenna is shortened from that length, losses in the feed system (even what Hams consider a good one) climb rapidly. 88 feet is just too short for an 80 meters antenna, because as you see above it is at the limit of what most tuners will match. It also places most tuners at their power limit at a few hundred watts of applied power.

The optimum length for a multiband dipole is near 1/2 wl on the LOWEST band, and the optimum open-wire feedline length is any odd multiple of 1/8th wavelength on the lowest band. This means an optimum 80-meter dipole would be about 125ft long, and the feedline would be 25-30ft, 75-90ft, or 125-150ft long. The longer the feeder, the more likely you are to having to trim it for optimum tuner performance.

Most tuners like to see impedances HIGHER than 50 ohms, and inductive loads at low impedances. Pi's and L's are NOT a solution to matching problems. They actually are significantly more restricted in matching range than a conventional T using the same general style and quality components!

Feedline voltage is also a good way to estimate wet-weather performance of "window" ladder line. If voltages are more than 1000 volts RMS at 100 watts, operation in wet weather will certainly cause tuning or loss problems. Use TLA and other tools as a way to plan antennas. Remember, there are more important things than pattern! A good pattern is useless if you cannot feed power to the antenna,

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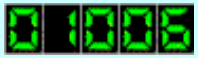
This page has been viewed  times since creation on July 25, 2004. This entire web site now averages over 1.5 million hits per year, and the entire web site now averages about 500 **different** visitors each day.

End-fed vertical j-pole and horizontal zepp

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Related Pages:

Verticals and baluns

The rule:

There is a rule that applies to end-fed antennas that cannot be broken. The rule is:

Current flowing into the antenna must be equaled by current flowing into a ground or counterpoise.

This means if you have any type of end-fed antenna without a radial or counterpoise system, the outside of the coax shield must carry an equal amount of current to that flowing into the antenna! This is true no matter how many series traps or choking devices you add at the feedpoint.

End-fed Myths

There are many incorrect ideas and claims surrounding end-fed vertical antennas (like the I-Max 2000), end fed horizontal antennas, Zepp, and J-pole antennas. All of the above antennas are actually quite similar. The J-pole is closer to the original Zepp antenna (Zeppelin antenna) than the bent antenna we commonly call a Zepp. I'll examine the most common and troublesome myth or false claim, that such antennas do not require a ground or will not have troublesome ground currents flowing over conductors into the station ground.

Some myths have been perpetuated or re-enforced by use of incomplete models. The most common modeling mistake is omitting the feedline or, if included, the coupling device (tuning or matching unit) properties are omitted. The major problem often isn't feedline radiation, it is common mode currents from the feedline. This article will explore why those currents occur, and give worse and best case examples of things affecting common mode currents.

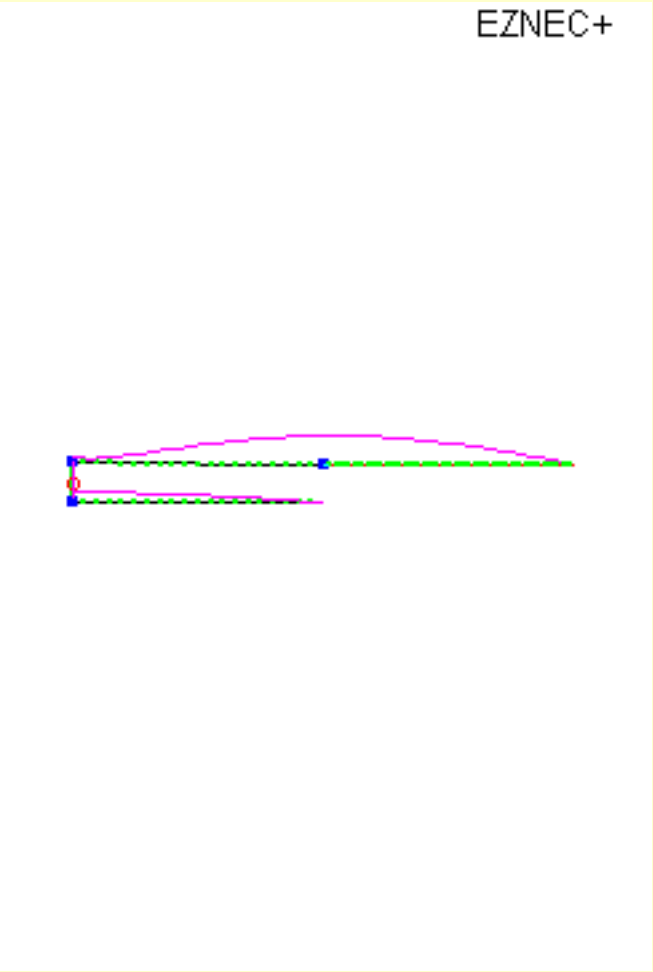
Origination of J-pole and Zepp Antenna

This form of antenna originated as a practical solution to feeding a wire antenna trailing behind a dirigible. This type of antenna is a ***Marconi antenna***, and it requires a counterpoise of some form.

(Marconi, by definition, means an antenna that is excited against a counterpoise or ground. An end-fed antenna is really a Marconi antenna, since it is impossible to force current into the end of an antenna without equal current flowing in some form of counterpoise.)

One way to create an effective ground system or counterpoise would be to use metal structural parts of the airship. This could create a problem if an RF-induced arc occurred where a mixture of hydrogen and oxygen was present. Running high RF current levels around inside a large bag of hydrogen with a pod of people hanging below was considered poor engineering.

The solution was use of a trailing counterpoise wire. With a second trailing wire was a counterpoise, the antenna system cold be floated from the airship frame. By floating the antenna and ground from the structure, most of the RF currents would be confined to the trailing wire and trailing counterpoise. The initial trailing wire antenna was 1/2 wavelength long, and the counterpoise 1/4 wl long. This set up a high impedance antenna feed working against a low-impedance counterpoise. Most of the current was in the antenna, as seen in this model:



Note the current curves in this model. Maximum current in the counterpoise wire is much less than the maximum current in the 1/2 wave section. Currents at 500 watts are:

- .71a into the 1/2 wl wire
- .71a into the counterpoise wire
- 2.32a maximum at the middle of the 1/2 wl wire

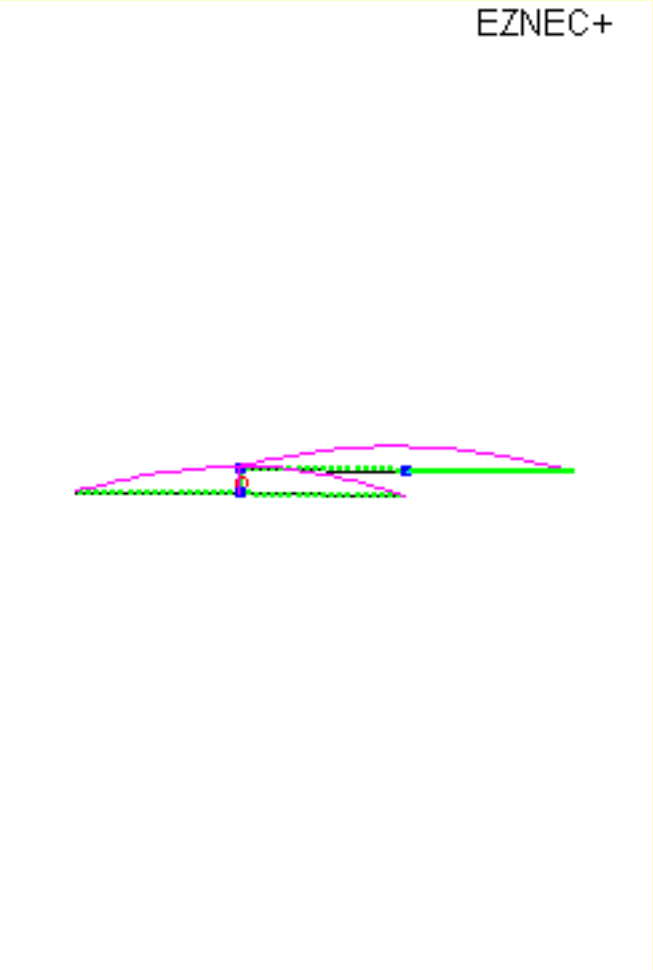
(Pattern is similar to a 1/2 wl dipole at the same height.)

Notice how currents fit the rule I mentioned at the start of this article. At the feedpoint the counterpoise wire, because there is no other path, carries exactly the same current as the antenna wire.

This model, like the models almost everyone uses, has an omission.

The flaw in this model is the same modeling omission K6STI included in his Zepp article, and the same omission repeated in "A Tiger by the Tail". The models do not include real-world effects of the antenna coupling system! All three examples feed the antenna with a ***perfect*** ground independent current source. That source that has no path to ground or to the station wiring.

When you include an imperfect source, the tiger turns around and bites you squarely on the rump. Here is what happens when a worse-case ground path is added:



In this case, I added a 1/4 wl wire to represent a low-impedance common mode path from the counterpoise side of the antenna connection. We see ground currents almost equal maximum antenna current!

In reality, most antennas will not be fed this poorly. A typical GOOD tuner has a few thousand ohms common-mode impedance. (There isn't any way around this, it is an impedance limitation inherent in the devices. You especially can't get around the problem by moving a balun to the input of a tuner!)

500-watt system currents when a typical maximum common-mode impedance is added are:

- .71a flows into end of antenna
- . 1a into ground
- .61a into antenna counterpoise

It is the .1a that causes all the problems. If you do not have a very good low-impedance RF ground on your tuner, this significant 100mA of unwanted RF current will flow into you, the power line,

the power supply, the microphone, and everything else connected to the radio and tuner! Of course the current will divide by the ratio of impedances, but this is precisely why, contrary to what other authors might claim, RF problems are common with Zepps. As a matter of fact, K9STH (a respected person on QRZ forums) uses a Zepp in an non-ideal configuration and has become a big proponent of extraordinary station RF grounds. This isn't meant to criticize anyone, but it is necessary to call attention to the problem of model sources being perfect while real-world systems have limitations.

The fact is, ***if an antenna is fed properly there is absolutely no need for a station RF ground! The station RF ground is a band-aid for poor antenna system design.***

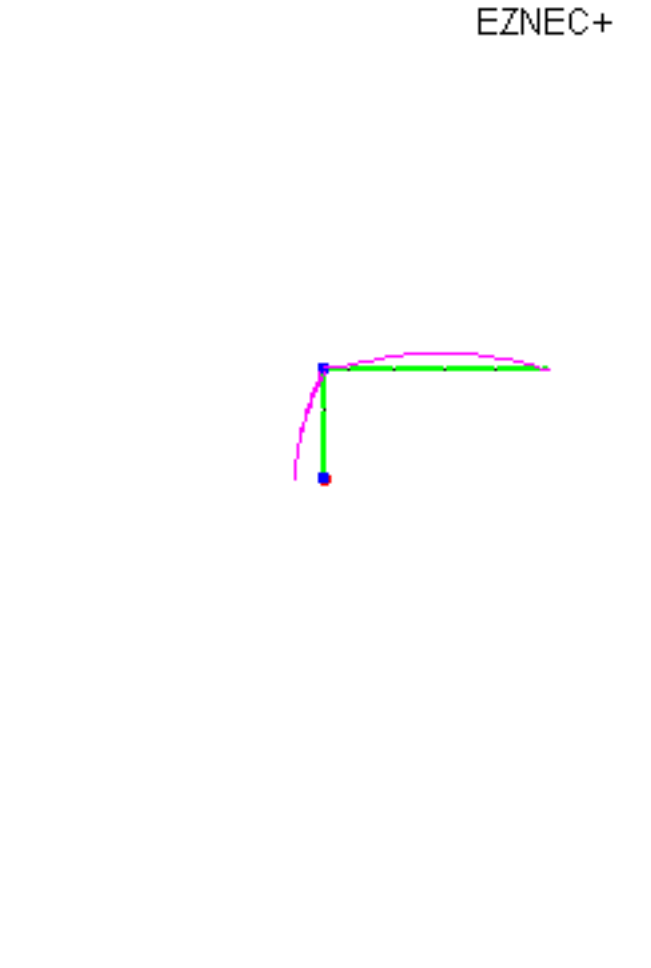
If an antenna is modeled properly to include feedlines and tuners, you will see problems that idea-source models overlook! Conclusions about system behavior change when tuners and transmission lines are included in models.

The Next Zepp Step

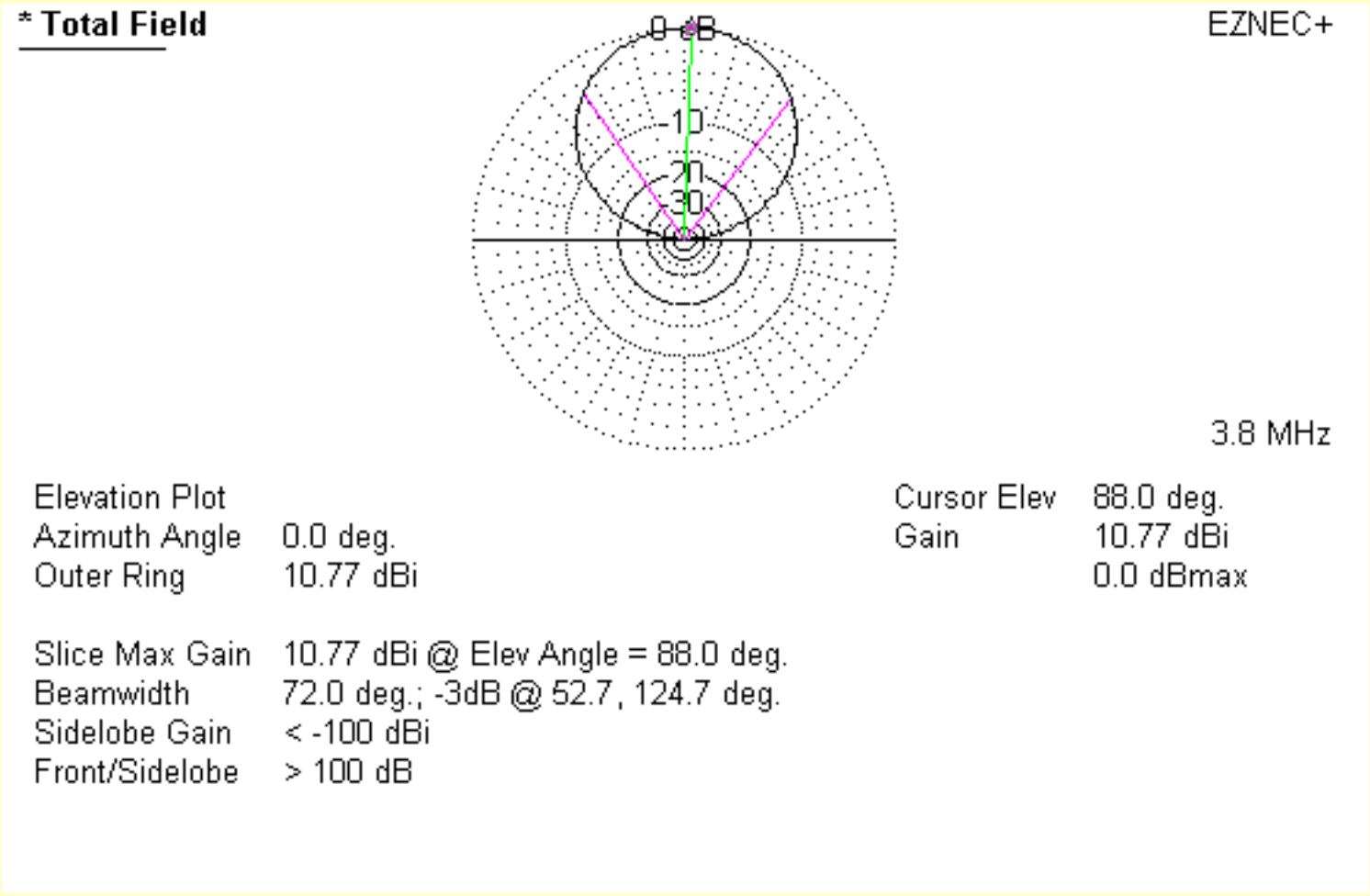
Let's look at the next evolution of the Zepp (this applies to end-fed verticals like the I-max 2000 although I will cover end-fed 5/8th wave verticals in detail at a later time). In this case we will model a 1/2 wl end-fed antenna fed with a 1/4 wl transmission line (a conventional year-2000 Zepp antenna), and see what happens when the 1/4 wl feedline is tied to ground through the common-mode impedance of a typical better-grade antenna tuner.

The ideal case for Zepp or J-pole (they are the same except for the bend in the Zepp at the feed) is when the radiator is very thin, exactly 1/2 wl long, and fed with a parallel conductor stub that is exactly 1/4 wl long.

Here is the antenna view:



Ideal antenna shows no pattern distortion:

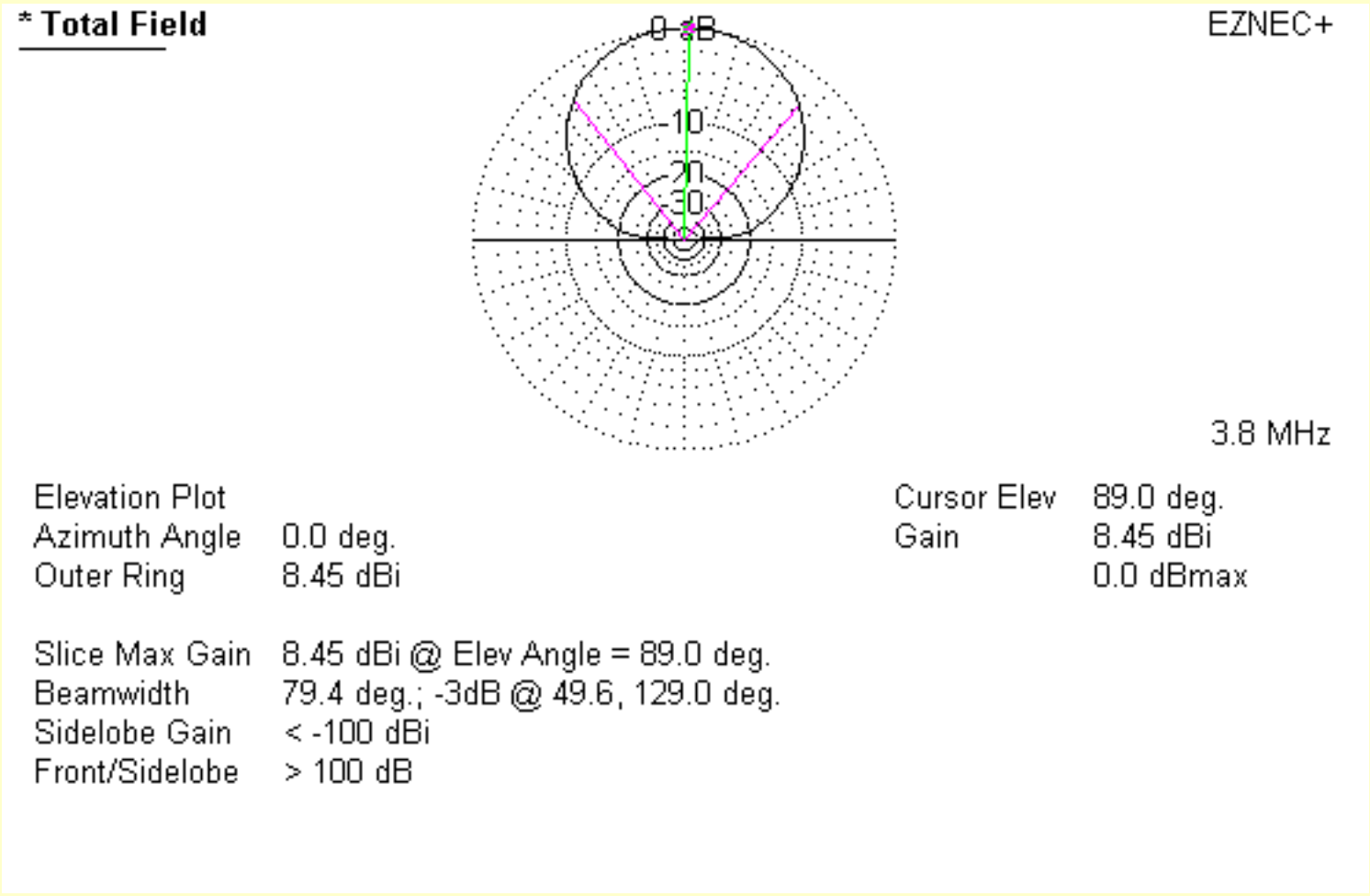


Although high feedline currents appear, they are closely balanced and out-of-phase. This is why both W4RNL and K6STI concluded feedline radiation is not a problem, and I agree ***for this specific perfect case!***

The real world adds a few variables, however. Real-world sources are not ground independent, real-world antennas are not always 1/2 wl long, and real-world feedlines are not 1/4 wl long.

Let's explore what happens when a the real world enters the perfect model!

Adding a worse-case ground path to a perfect antenna, we have:



Now we see a slight flattening of the bottom of the pattern. While the pattern does not change much, we have the following currents at 500 watts:

- 6.1 amperes feedline wire 1
- 5.35 amperes feedline wire 2
- .75 amperes ground path

Even with the small pattern error, we now has a very significant .75 amperes flowing to earth through the station equipment and perhaps eventually the operator. This is with an IDEAL antenna, and a normal worse-case antenna feed.

What is a worse case feed? The common J-pole antenna fed with coax! .75 amperes would represent the undesired shield currents and supporting pipe or mast current total when a J-pole is fed with coax.

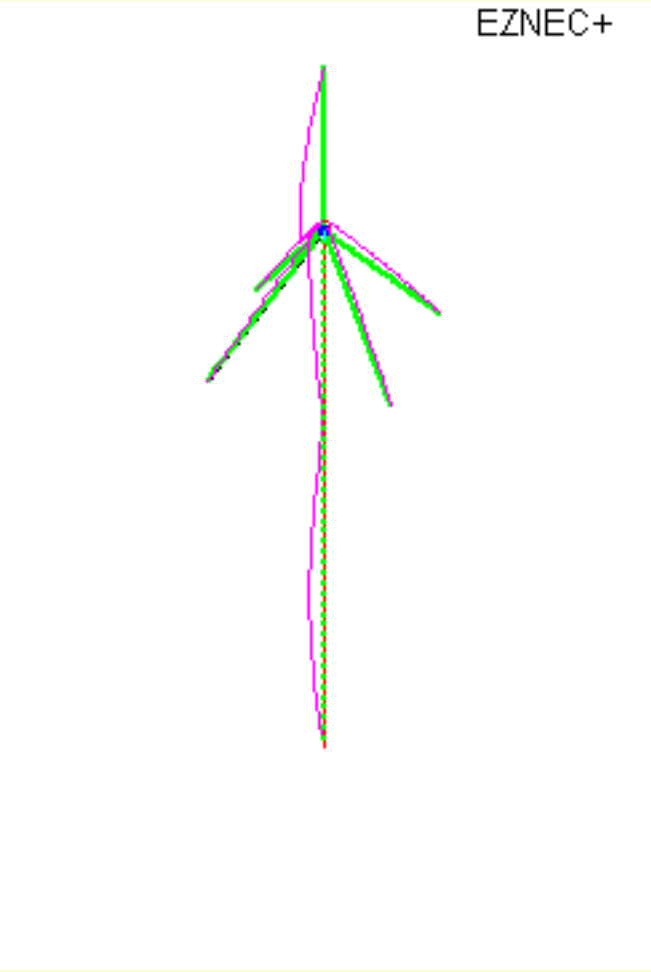
This problem gets worse when the Zepp feedline is not a 1/4 wl long, and the antenna is not 1/2 wl long. The I-MAX 2000 CB vertical is one example. The antenna has a random feedline (your coaxial feedline and mast) as a counterpoise, and is not a 1/2 wl end-fed vertical! This is why so many people complain about RF feedback with Zepp and end-fed 1/2 wl antennas.

If time permits I'll add examples of worse length combinations, but this example of a best length proves common-mode currents are an issue with a perfect Zepp connected to a real tuner. This is why people with Zepp are generally people who believe good RF grounds are required. They are correct in that belief. if you have a poor antenna design you need a good RF ground at the station!

Ground Plane and J-Pole Problems and Patterns

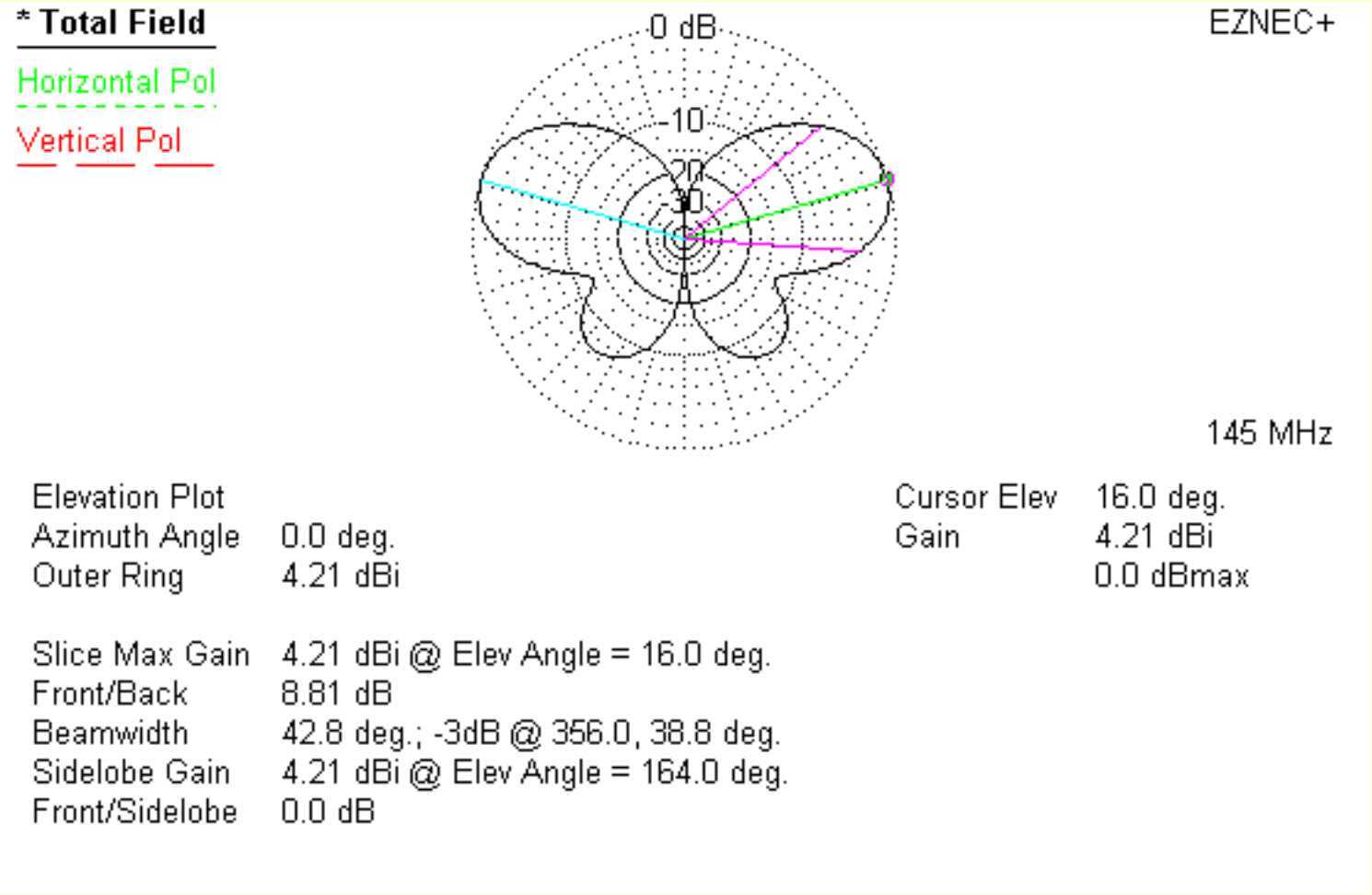
The J-pole is a worse-case situation for examining unwanted feedline radiation. In the case of the J-pole without radials, all of the common mode current created by the poor feed system flows over the coax shield and mast. The mast and coax is generally vertical and as high and clear as the actual thing we think is the antenna, and so unwanted radiation from the feedline is much more apparent than in the low-band Zepp where both feedline and antenna are near earth.

We will use this as a reference. Here is the pattern of a ***1/4 wl groundplane antenna*** without proper feedline decoupling:



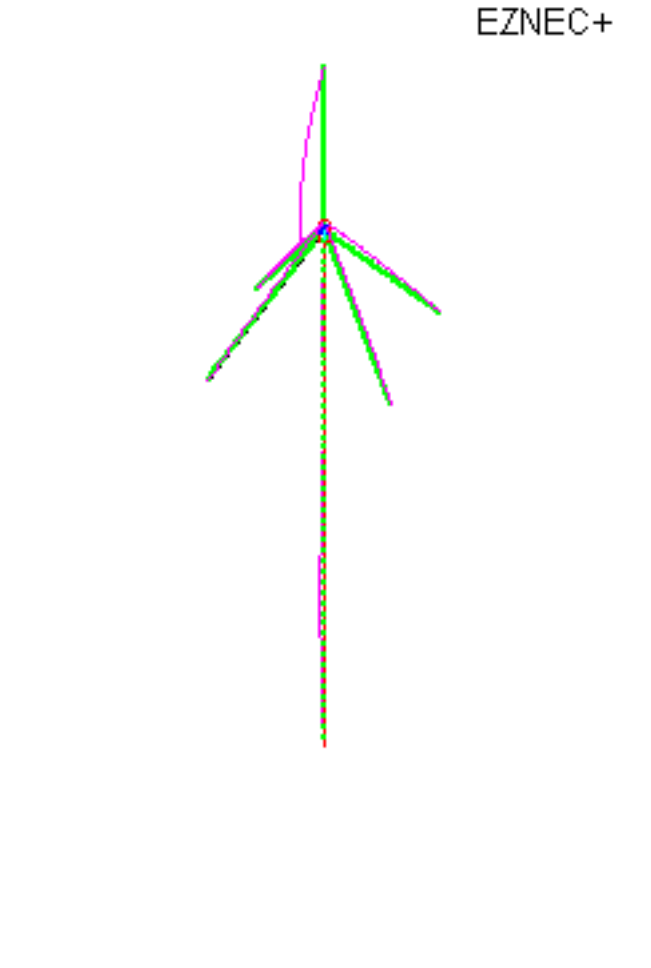
You can see some current exists on the feedline. The amount of current is actually significant, almost 1 ampere compared to 2.5 amperes in the antenna (500 watts applied)! A 1/4 wave groundplane with four radials actually needs a balun or some other form of feedline decoupling.

The pattern in freespace of this antenna is:

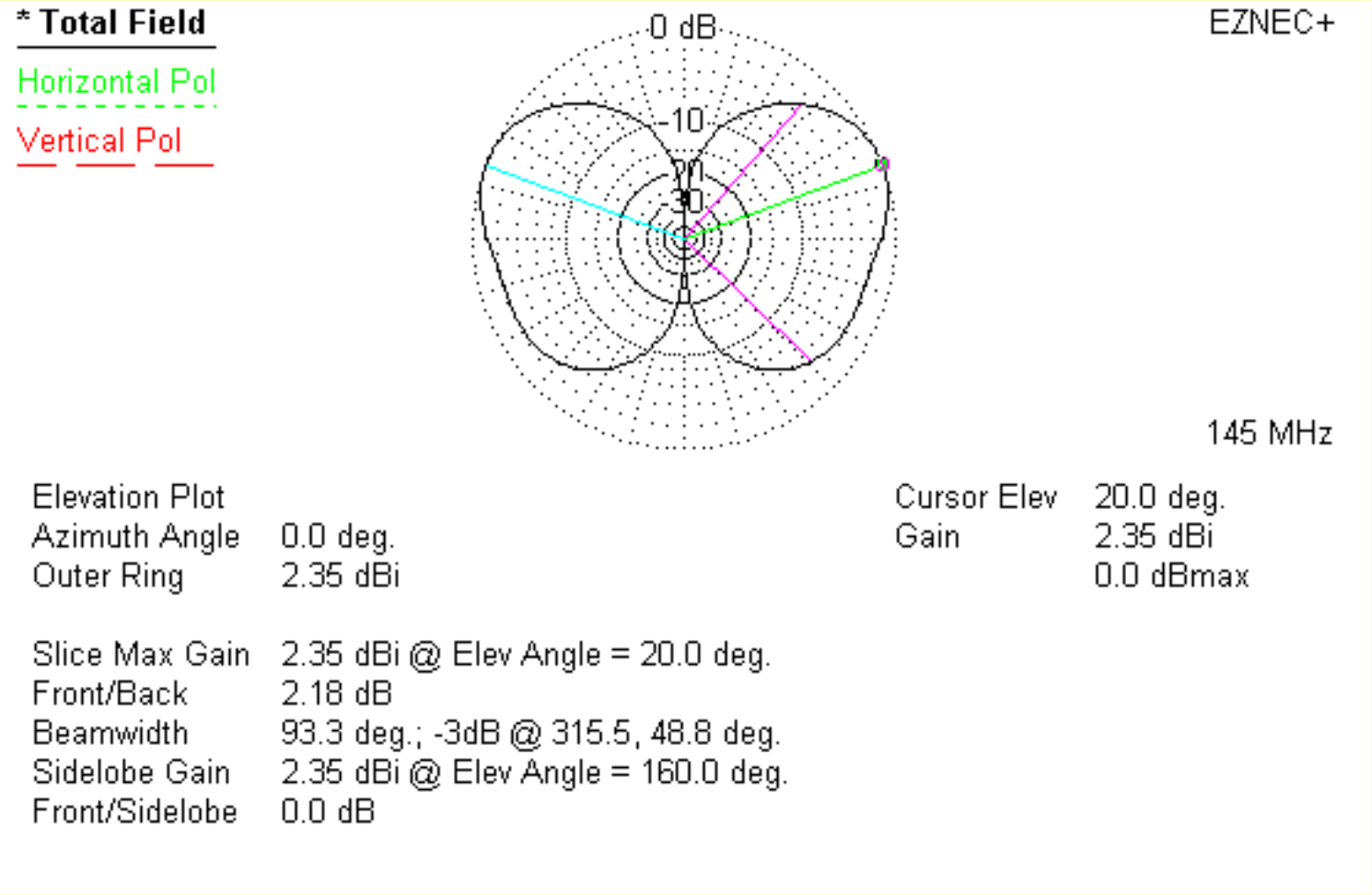


The pattern distortion is caused by feedline common-mode current. If you think this is bad, imagine what happens with an end-fed antenna (even a 1/2 wave antenna) and NO radials! In that case all of the antenna current flows over the feedline shield! Many antenna designs actually use the feedline and mast radiation that others dismiss as "insignificant" to increase antenna gain. In some cases, the antenna designers really don't even understand what they did to create a "magic" antenna.

Here is the same 1/4 wl groundplane with a feedpoint balun with a common-mode impedance of 3000 j3000:

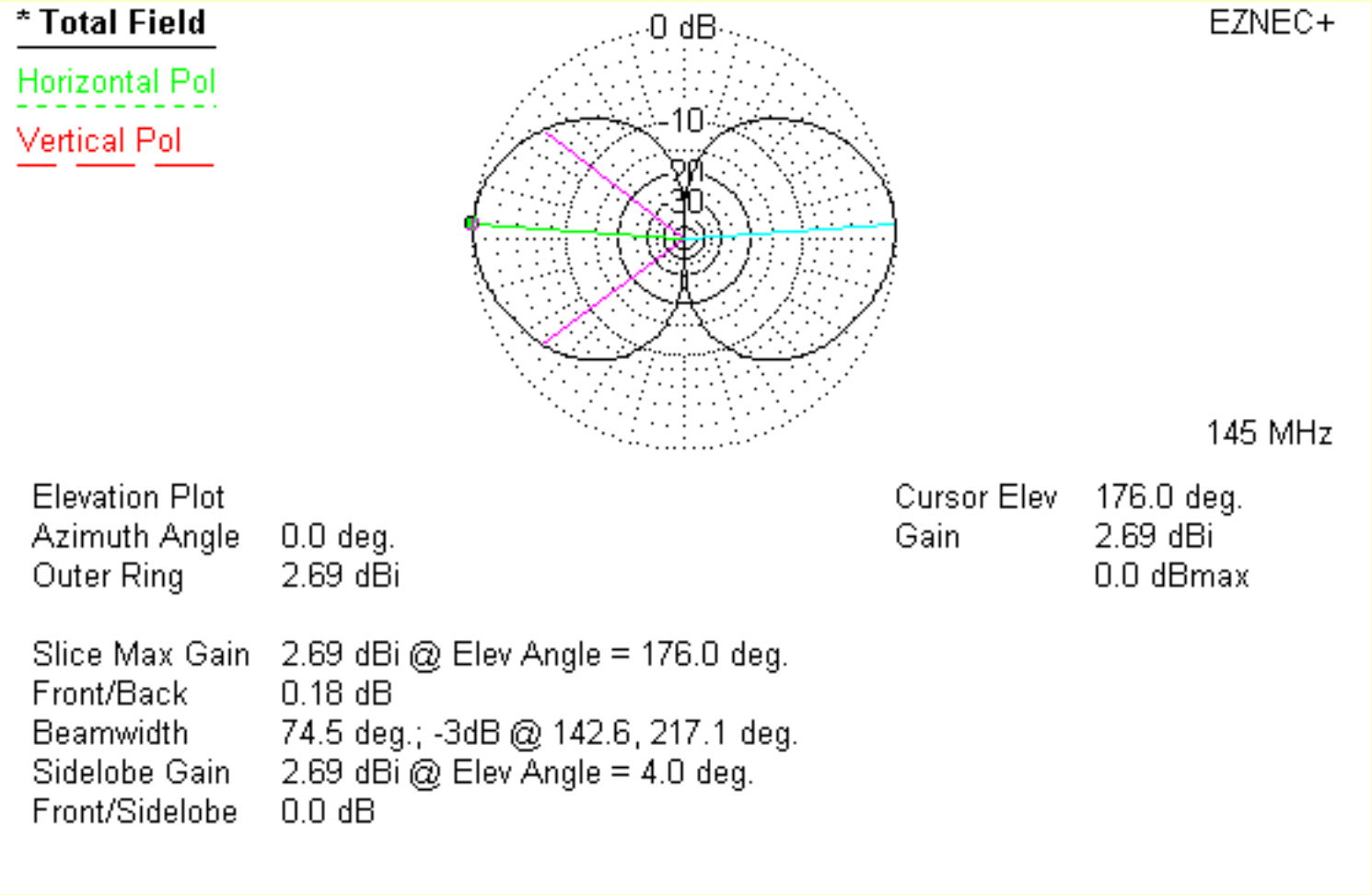


In this case feedline currents are under .4 amperes, and antenna current increases to 3 amperes. pattern is:



See how much cleaner the pattern is? The pattern is much cleaner and more like what we expect from a groundplane far above earth. We still don't have a good system, because the balun needs a ground on the "unbalanced" side to be effective. (This is why I recommend grounding the unbalanced side of baluns to antenna booms of Yagi antennas. That coil of wire hanging in the air might not be enough without a ground!)

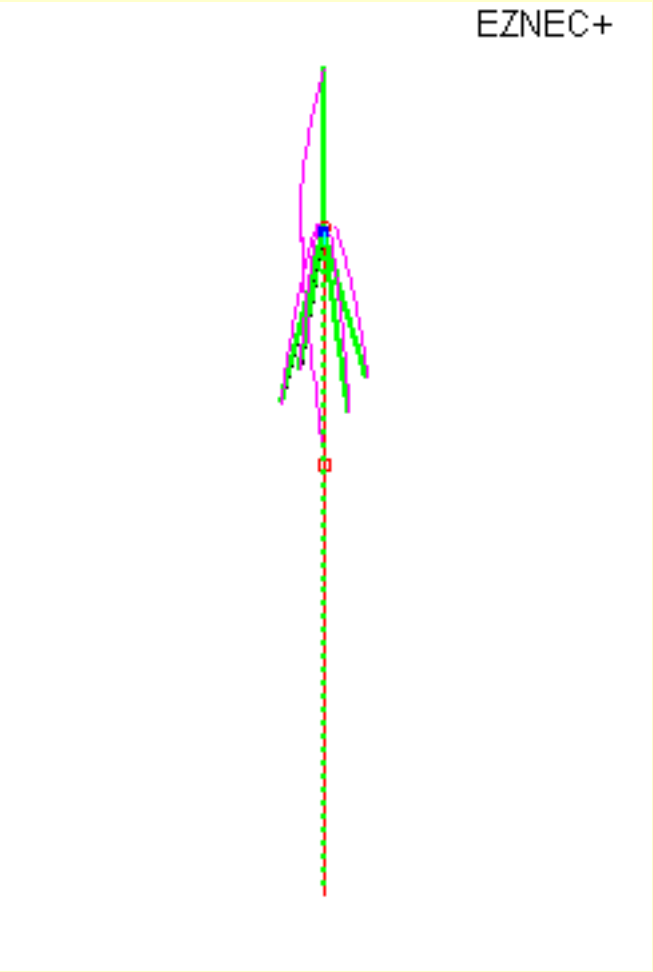
With a perfect balun and optimum de-coupling (a few extra radials below the balun or an extra-ordinary balun), the pattern looks like this:



This is the ideal pattern of a 1/4 wl groundplane, and it takes extraordinary care to obtain this pattern. Many people think they have this pattern, because they model antennas ***without*** *the feedline attached!*

To obtain this pattern I had to:

- Use a good balun just below the radial tip level
- Ground the feedline shortly below the balun to a small groundplane
- Slope the radials downwards more, to a 75 degree angle.



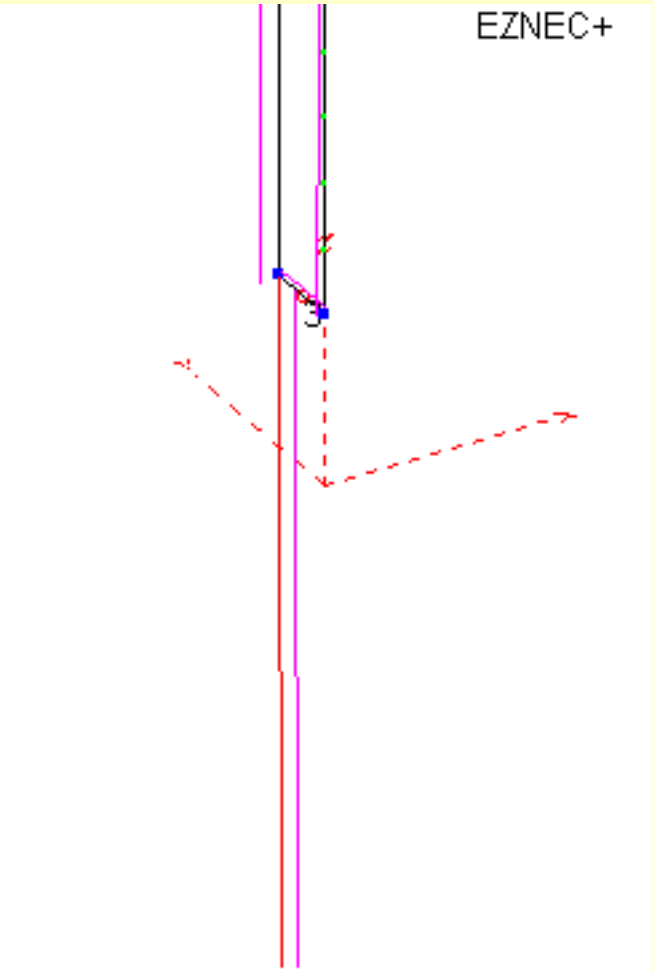
It is easily possible to do this in the real world. Instead, we have numerous articles that ignore the feedline and assume we don't have a problem with common mode current upsetting antenna pattern.

Worse yet, some articles claim the groundplane antenna only needs one or two radials!! Why does this happen? Because the person who made that assumption never modeled the mast or feedline. They eliminated the real-world problem through an omission in the model.

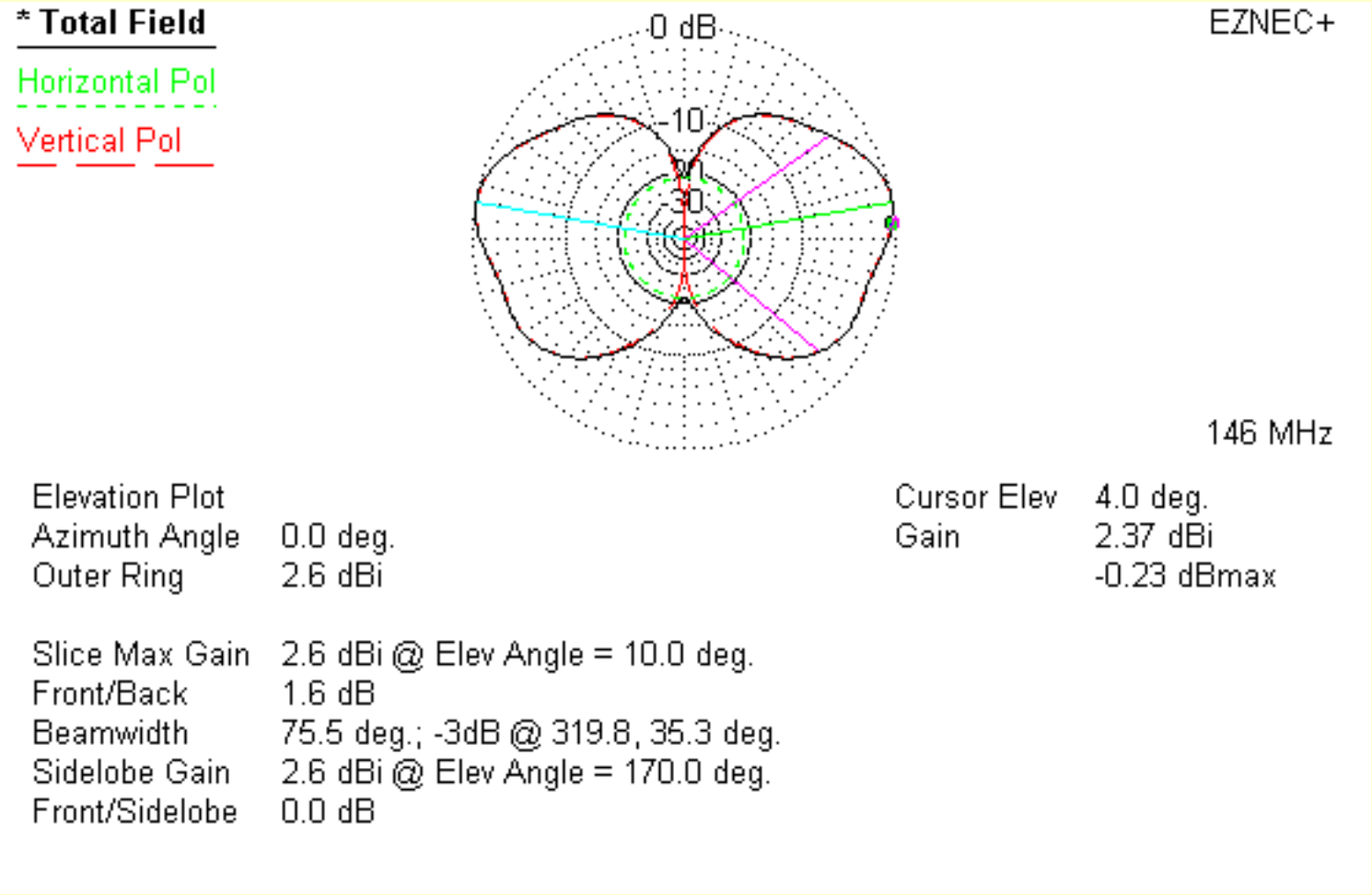
If an antenna model does not include the feedline, losses in the feedline, or imperfections in the source it much more likely than not is a flawed model. The real-world antenna often will be much different if the model ignores feedline loss and feed system source or matching system problems. Short dipoles are one example.

The J-pole and other end-fed Hertz antennas as prime examples of antenna with severe common mode current problems. The coax shield has to be at zero volts potential and have exactly equal and opposite currents flowing into and out of the load and source, otherwise the feedline radiates. Feedline length or weather changes that affect feedline moisture between the outer jacket and the support for the feedline will often change SWR. The severe common-mode feedline problems of end-fed 1/2 wave antennas is why some people swear by them, and other people swear at them.

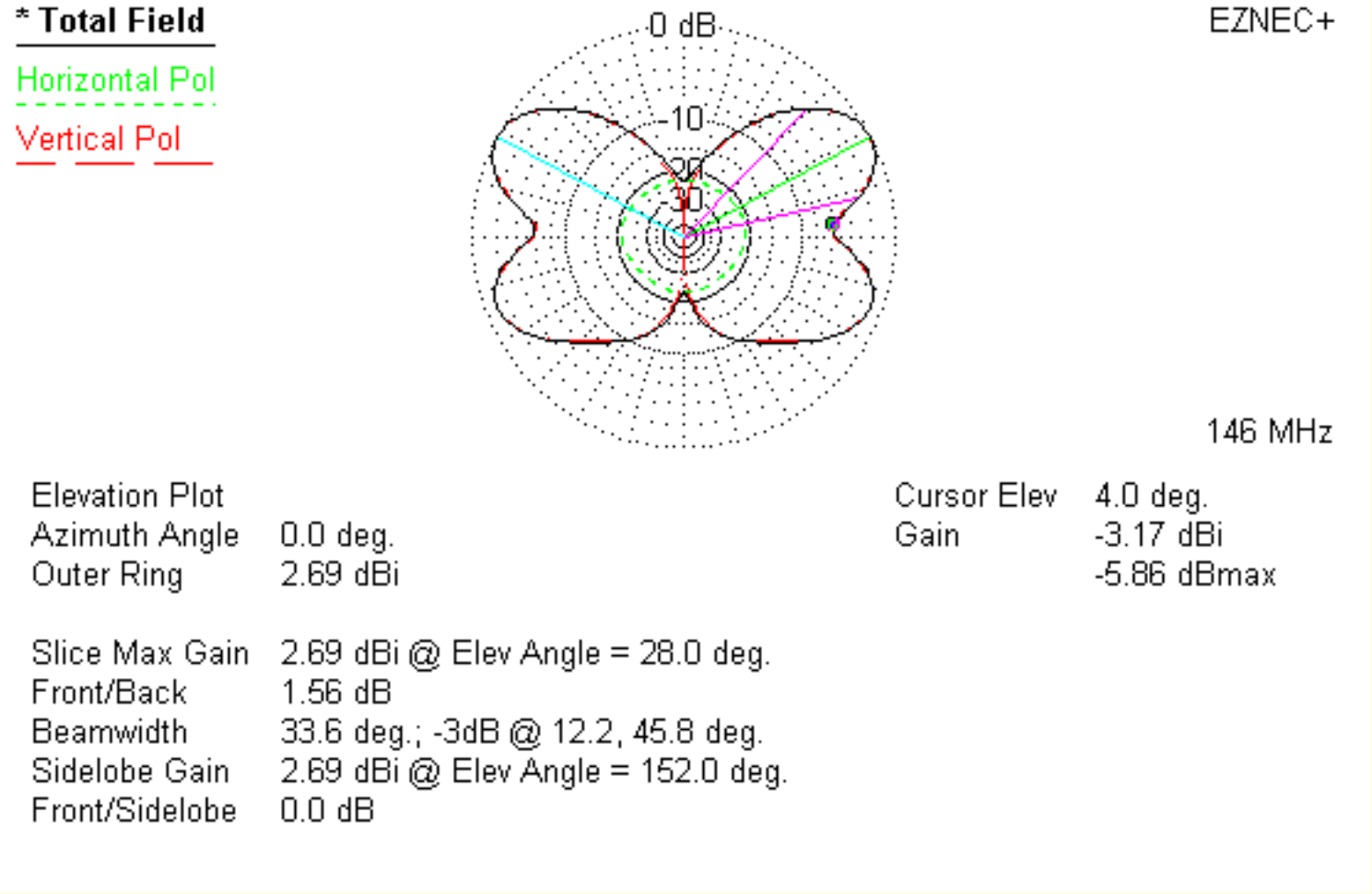
Here is a zoom of a correct model of a J-pole with a vertical feedline and/or mast attached.



You'll see the feedline or mast grounds directly to what everyone assumes is a "zero voltage" point. This is the electrical equivalent of any J-pole with the coax connected in series with the feedpoint, and the shorter leg connected to the shield. The shield can be connected to any supporting mast with any change in results. Here is the resulting pattern:

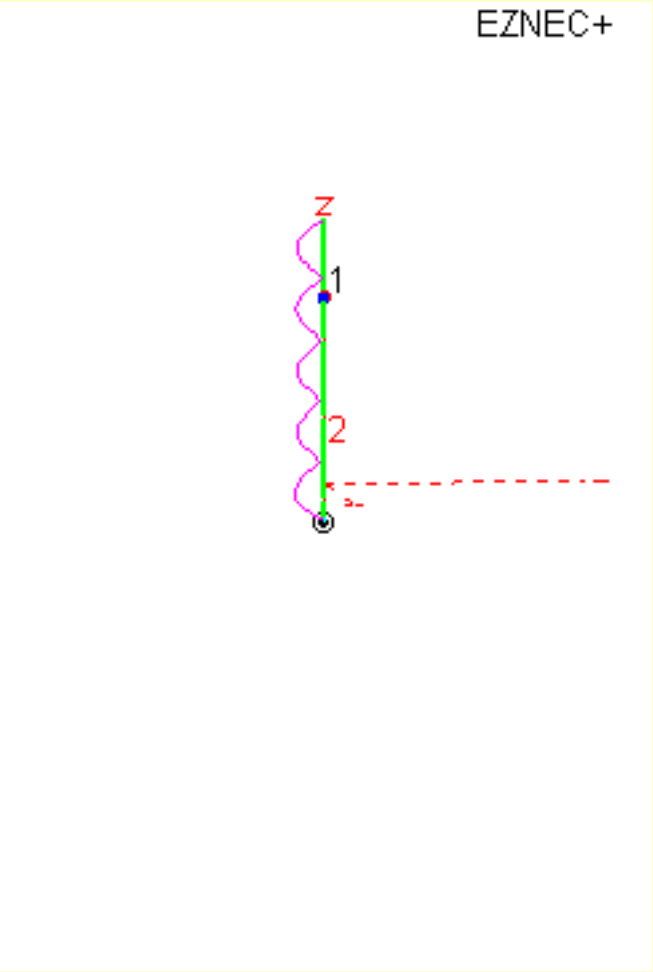


The gain is now 2.37 dB at 4 degrees elevation compared to 2.69 dB for the 1/4wl groundplane. This isn't the worse feed arrangement....it is actually the best for the J-pole! Here is the pattern with the feedpoint reversed, the shield is connected to the longer element, and the center conductor to the short element:



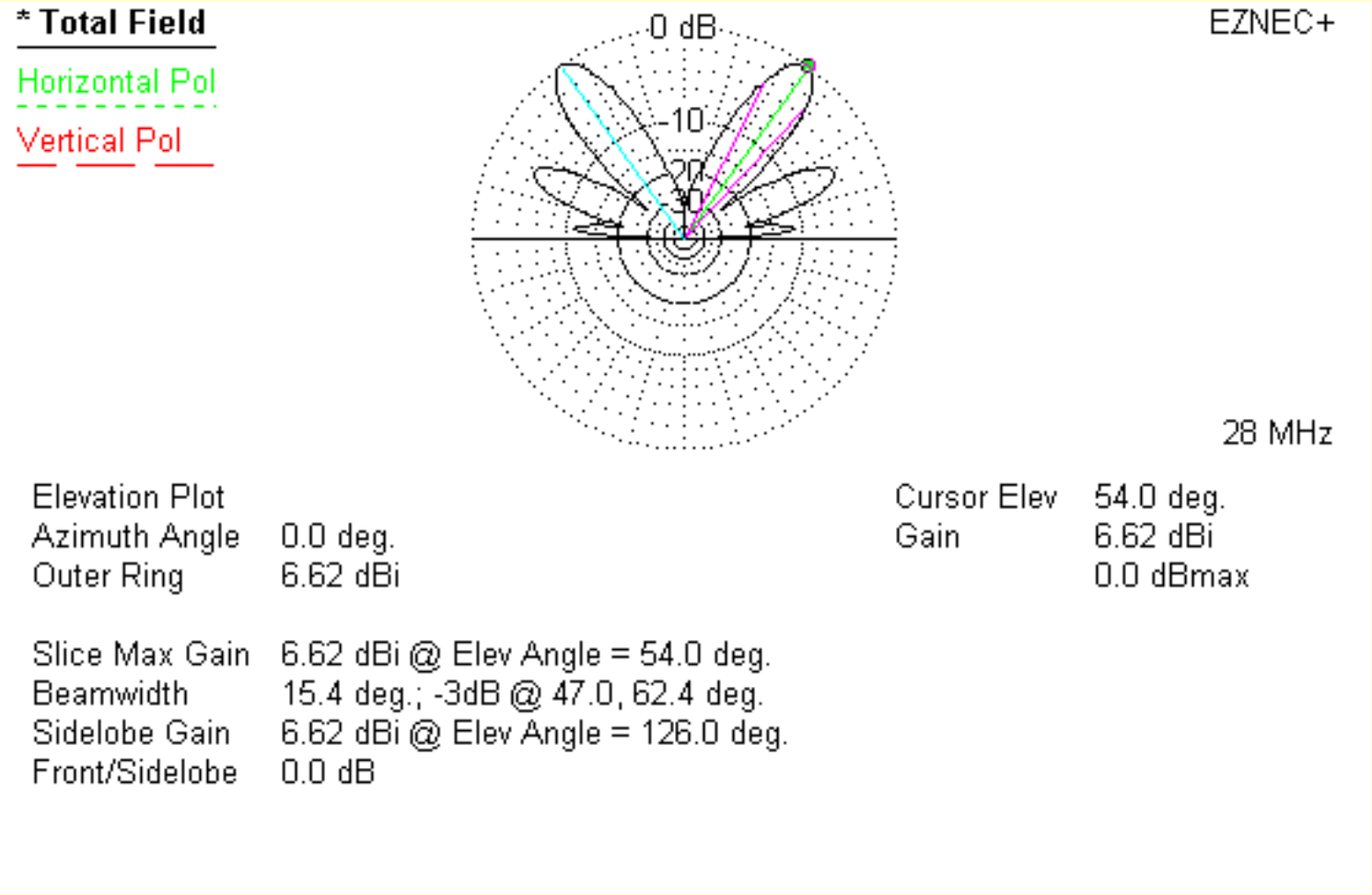
Notice the low-angle gain dropped about 5dB with just a simple reverse of feedline connections! If I didn't model the feedline, the model would never show this problem. In ALL cases, the SWR stays near 1:1, yet gain at low angles changes 5dB!

The following model is an I-Max 2000 vertical with a vertical feedline. In this case I picked the worse feedline length:



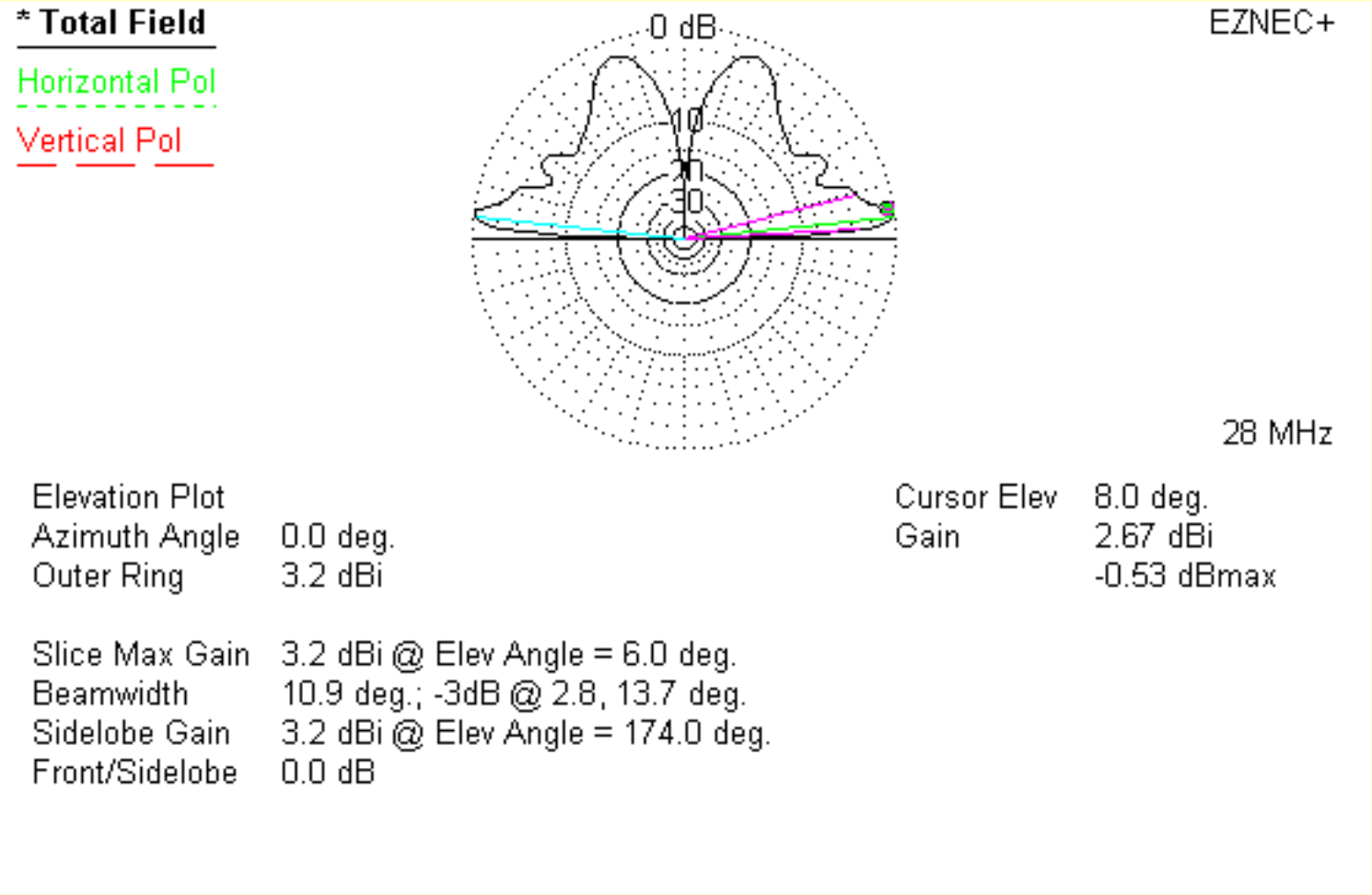
Feedline current is 100% of antenna current. This illustrates why so many people complain about SWR problems and RF in the shack with end-fed verticals like the I-MAX 2000!

Here is the pattern of an antenna that copies the I-MAX dimensions and feed system:



This is a NEGATIVE gain antenna at low angles. A 1/4wl groundplane would seriously out-talk the I-MAX 2000 or any other 5/8th wl antenna that does not have a large groundplane.

Even if we use the optimum feedline and mast length, here is the very best the end-fed antenna will do:



In this case we now have 2.67 dBi, which is actually a little less than a 1/4wl groundplane will do! The severe common-mode mast and feedline currents make "no-radial" verticals extremely sensitive to mounting height, mounting structure, feedline length, and grounding. This is NOT normal for antennas, it is a sign of a design problem.

Summary End-feds Without Grounds

ANY END-FED ANTENNA REQUIRES A LARGE GROUNDPLANE OR OTHER EXTRAORDINARY ISOLATION METHOD OR METHODS TO PREVENT FEEDLINE OR MAST COMMON MODE CURRENTS!

This is true for 5/8th waves, Zepp antennas, R7's, R5's, or even common J-poles. End-feeding antennas is bad news unless you have a large well-established ground at the feedpoint. Even 1/4wl groundplanes have common mode problems.

Some manufacturers have wised-up.

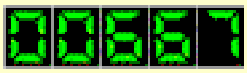
Cushcraft, in their Ringo-Ranger, eventually added a separate additional groundplane below the antenna to tame the wild common-mode currents of the Ringo. Even that solution is barely

acceptable.

The Isopole antenna used multiple sleeve sections to decouple the feedline, and it probably was one of the best antennas available for immunity to feedline coupling problems.

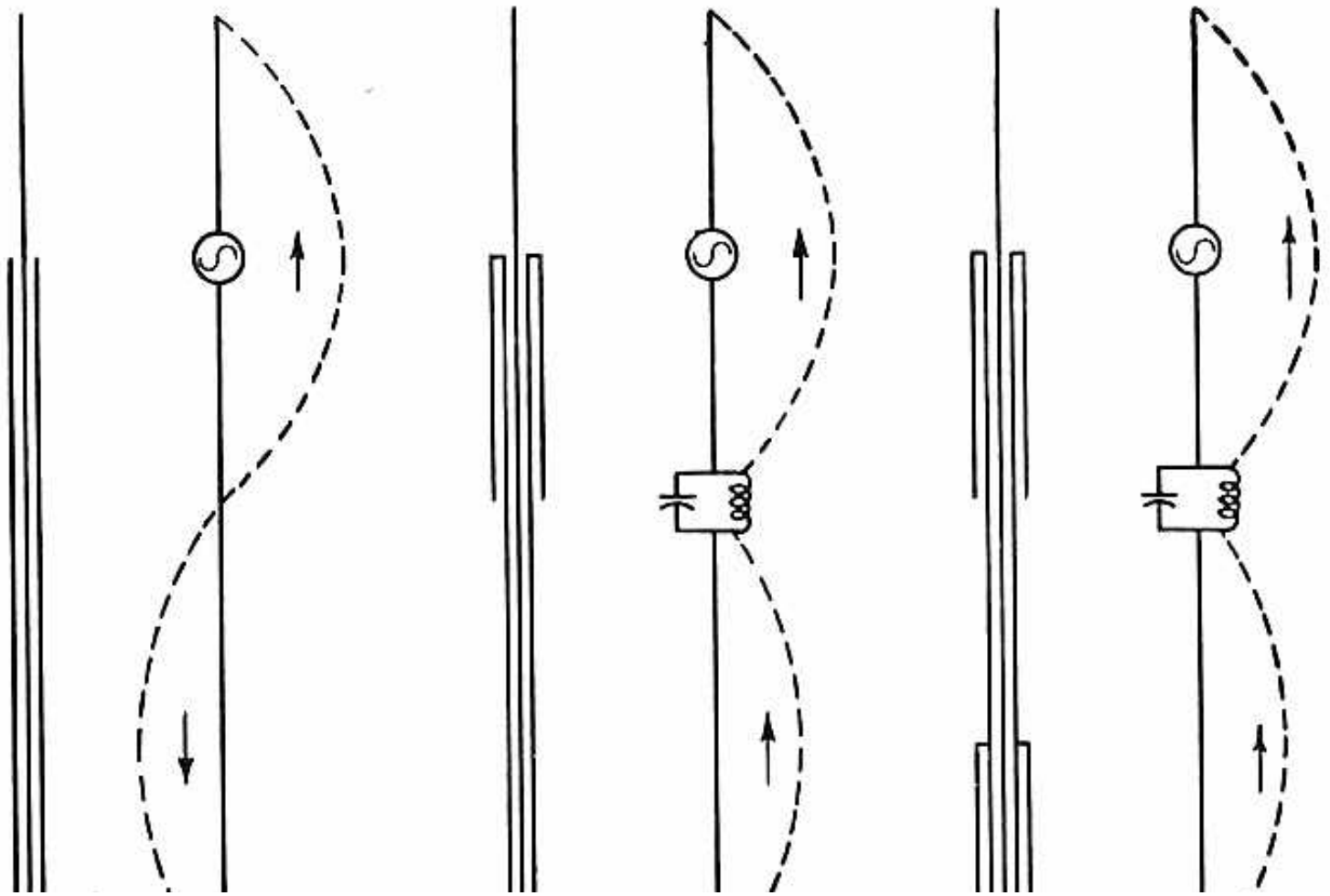
This problem gets worse when the element is 5/8th wave long. Think of that when you read claims of "no-radial" CB antennas with "3dB gain" and a low wave angle. They actually have negative gain at desired DX angles over a conventional 1/4-wave groundplane! Instead concentrating your signal towards the neighbor's TV set or an airplane flying overhead. They cause your entire antenna system to be critical for feedline grounding, routing and length and even allow moisture on the feedline jacket to change performance of the system!

If you find this page interesting, please pass it along. It's time to get the feedline into models!

This page has  **hits since creation on July 31, 2004**

http://www.w8ji.com/Skirt_feed.jpg

Two or even three skirts may be added to the mast as shown in Fig. 22-5*c*, but reduced excitation of the lower sections diminishes the effectiveness of each additio skirt.[6] Thus, while the multiple-skirt coaxial antenna resembles a collinear arra in-phase half-wave elements, its gain is not as great. Typical gain values are ta lated in Table 22-2, but since no test data are available for this type of antenna, values given are merely engineering estimates.



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http://www.w8ji.com/Skirt_feed.jpg

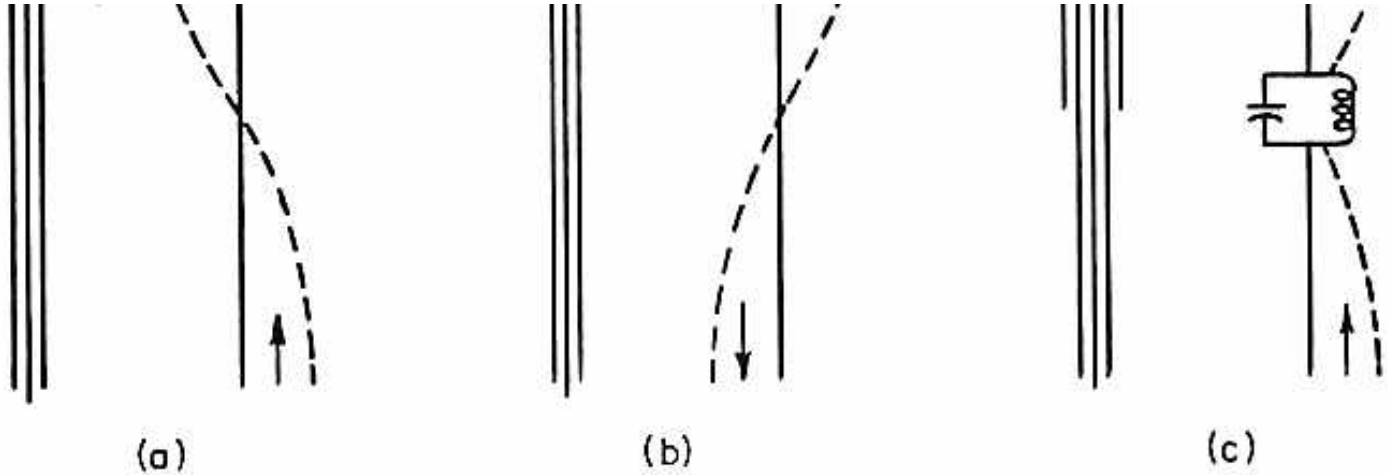

FIG. 22-5. Evolution of multiple-skirt coaxial antenna.

The relatively large diameter-to-length ratio of a practical skirt produces an effect which requires that the exterior length of the skirt be reduced by a factor of 0.8 to 0.9 and the interior length by from 0.95 to 0.98.[6] This is accomplished cutting the skirt to the proper exterior length and inserting a dielectric slug in skirt to increase its electrical length.

http://www.w8ji.com/Skirt_feed.jpg (2 of 2) [9/6/2004 9:26:54 PM]

diversity receiver and transmission

[Home]

Diversity Receivers

We sometimes are told a quad antenna, by virtue of the vertical side and horizontal top and bottom wires, is "diversity". That just isn't true at all. Neither is an intentional mix of vertical and horizontal reception from one antenna (like a Carolina Windom), or even two separate horizontal and vertical antennas when combined into one receiver signal input. I've actually never even found it to work very well even in stereo or with a voting system, as long as the antennas are within a wavelength or two of each other.

Despite what we are sometimes told, we can't simply adjust two receivers to the same basic frequency, pipe one output into each ear, and have diversity. This is especially true with unlocked receivers that are even a fraction of a Hz different in frequency.

There are articles describing locking the dials of the FT1000D so the sub-receiver tracks the main receiver. This does work in the FT1000, because the oscillators share a common time base. The main problem with the FT1000 is the second receiver isn't a "good" receiver.

Testing For Diversity Capability

You can test your system for correct phase-lock by tuning in a carrier (like WWV) and mixing the two outputs together. When audio levels are equal, you should hear no warble or vibrato in the tone. You should hear something between a full or peak that stays perfectly locked. If your system fails this test, it will deduct greatly from possible diversity advantages. Click here for a failed test. Click here for a passed test.

Remember any warble or slow fade variation indicates the receiver is unacceptable.

True vs. Stereo Diversity

I use a loose form of diversity reception on 160 and 80 meters I call "stereo diversity". It really isn't true diversity where the receivers vote and the best S/N ratio captures the audio output, it is a system that requires your "brain" to do the voting.

I've found this system very good for substantial improvements in readability of noise-floor signals. The difference can be worth as much as a signal being nearly readability 5 (perfect) in stereo to readability 2 without. When a signal is marginal, it is all the difference in the world.

I accomplish stereo diversity by phase locking two separate receivers together so audio outputs are exactly in phase. The receivers are virtually identical, even to the point where I hand select crystal filters for equal group delay change over the filter passband. Every oscillator in the receiver is common to both receivers.

This allows me to successfully combine two antennas or antenna arrays that are several wavelengths apart into one system, a task otherwise impossible. Mixing two quiet widely-spaced antennas causes noise to appear smoother and more "hollow sounding", while the signal actually stands out more. Some of the recordings on my DX Sound Files page are in stereo. You can listen to this effect with stereo headphones by changing your computer's volume control settings to stereo or mono.

Transmitting

One common thought or claim is by transmitting both vertical and horizontal via skywave, we would have the best of two worlds. Worse yet, there is a common idea or belief we can build antennas producing two independent polarizations, and that resulting dual polarization will provide the best of two worlds and reduce transmitted signal fading. Several obvious flaws with this concept are outlined below.

Generating Two Polarizations

When speaking of polarization, we are talking about the direction of the imaginary flux lines in the electric field. When dealing with the far-field effect called EM radiation, ***the imaginary electric and magnetic flux lines are conveniently at right angles to each other***. While either could have been used for the reference standard of polarization, the electric field became the polarization reference.

The imaginary flux lines represent the force caused by any and all electric fields. They "exist" only at one angle in one small portion of space at any instant of time. We ***can not generate two polarizations at the same time at any reference point in space*** when broadcasting our signals, not with any antenna! The idea we can have dual-polarization transmissions probably comes from misunderstanding what antenna modeling programs are showing us, or a flawed or limited imagination causing an incorrect mental picture of what actually occurs with antennas.

Modeling programs only show two perfectly filtered views of the actual field. They do this out of

necessity, because the actual electric field is far too complex to display in its entirety. While we could see a slice of the field showing true polarization at any given angle and distance, I'm not aware of any commonly available programs that provide such useful information. A typical pattern display generally shows the response that would be observed through perfect vertical and horizontal filters. Modeling programs generally do not tell us the phase relationship between the intensities we see displayed, so we have no idea what the actual polarization is.

The bottom line is this....we don't know, when looking at the display, what the actual polarization is unless it is 100% vertical or 100% horizontal.

A Simple Tilted Dipole

Visualizing an actual antenna might help us picture an antenna pattern correctly, and understand what we commonly perceive incorrectly.

Imagine we have moved back a considerable distance from the center of an extremely high dipole that was installed tilted at 45-degrees. We move back from the dipole center without changing height, and observe the electric flux lines near us. To us, the distant antenna appears tilted at a 45-degree angle from our lower left to upper right. Each end of the antenna is exactly the same distance from us. In other words, this is a "side view" of a perfect sloping dipole.

If we could actually see flux lines near us representing the electric field, the lines would appear to parallel the distant antenna. Yet a view on a modeling program would show the field intensity of the electric field to be an exactly equal mix of vertical and horizontal fields!

Many of us would (incorrectly) describe this antenna as producing equal vertical and horizontal polarization in the direction where we view the antenna. The logical conclusion would probably be a vertical or a horizontal would respond equally well to that field, and that is correct for a perfect vertical or horizontal.

What we probably would fail to understand or visualize is the actual polarization. The peak response would be to a dipole antenna tilted 45-degrees in the same slope direction as the radiator, from lower left to upper right. Most important and most often missed is the simple fact that another dipole tilted 45-degrees opposite, from lower right to upper left (even though broadside towards the distant source) would have no response! It would be cross polarized, and response would be minimal.

This idea we have both polarizations is the root of the misunderstanding, and misunderstanding always seems to breed voodoo antenna claims and snake oil solutions. The most common false conclusion would be thinking sloped antennas reduce polarization related fading. The "equal-V-and-H-antenna" would be assumed to provide the best of two worlds, transmitting or receiving when either vertical or horizontal polarization or anything in between is required for an optimum signal.

Ionosphere Propagated Signals

Any distant signal arriving via the ionosphere is constantly changing in polarization. The ionosphere is a poorly aligned soup of ions, and that soup is constantly being stirred. It is not a flat perfectly aligned mirror. The ionosphere also provides multiple modes and paths for signals, particularly on frequencies well below the maximum usable frequency. The phase and level of the same signal arriving from each path constantly changes. Because of this, arriving signals tilt and rotate.

While there are some statistical odds that more time will be spent centered around one effective polarization than another, the fact remains that very little time is spent at one distinct polarization. The same effect holds true when transmitting.

Because of the random nature of polarization, the signal just as likely would be tilted 45-degrees left on on sloper as 45-degrees right. It is just as likely to fall into a cross-polarization null with the sloper as with any other angle of radiator, except one centered at the optimum tilt. We can easily see the idea sloped wires, Inverted L's, and even Windom antennas with "leaky" baluns reduce fading by providing "diversity" is pure rubbish. The same holds true for intentionally mixing two polarizations from two separate antennas, even if each is fed from separate amplifiers.

Statistically, we are actually MORE likely to have deep fades when we transmit with two very different systems than with one! The reason is simple, we excite the multiple paths better and increase multipath propagation. Since the phase delay is random and constantly changing, any attempt at circular or dual polarization would greatly increase fading when more than one path makes it to the receiver.

This is actually the reason 5/8th wavelength verticals fell out of use in broadcast work. The small high-angle lobe of the 5/8th wavelength antenna created severe deep fading and phase distortion at receivers in the fringe areas. A mix of horizontal and vertical antennas, even if properly phased to provide a rotating wave, would be even worse.

Reducing Fading

The best solution is to have two separate antennas, selecting the best antenna at any given instant of time for the path. This is true for both receiving and transmitting. Commercial sites do this by employing some form of voting, based either on signal-to-noise or absolute signal level.

Dealing with weak signals provides special problems. Good CW operators can copy code that is actually below noise floor. Because S/N ratio is near zero, noise detectors in a voting system would become overly complex and unreliable. Perhaps someone can develop a DSP system that allows

voting, but my attempts have been largely unsuccessful.

My solution is to use a stereo system with phase-locked receivers, and process the audio in my head. With antenna separations over a few wavelengths, the background "white noise" takes on a distinct hollow sound. Signals are easier to pick out, and the ability to copy CW below the noise floor is greatly enhanced. The end effect of this is reduced fading.

For transmitting, the only useful approach is having a variety of antennas available and picking the antenna generally more optimum for the particular distance, direction, and time of day. Without feedback from the receiver, it is all a guessing game. One thing I do know is that mixing my antennas directly never resulted in improved signal strengths or reduced fading in many dozens of blind tests.

How to Check Signals

[Home]

This page still under construction revised 1/8/2004!!

Related pages: Receiver tests Key Clicks Checking Transmitters

I noticed W2VJN has proposed using absolute signal level at a fixed test spacing for click measurements. While I agree with George's suggestion we establish a test method, I strongly disagree that a 15Hz BW signal level measurement of the peaks accumulated in dozens or hundreds of scans is any more useful that listening off-frequency on a known good receiver.

Some radios, like the FT-1000 series and other Yaesu radios, click on both make and break. Radios that do that are much more disruptive than radios that click only on make or break. The slope of the roll-off is also very important.

What we really need for click measurements is a peak and average power measurement on the adjacent frequency.

My local wintertime 350Hz BW noise (after preamplifier) compared to a sample of signals on one night was:

Noise -127dBm

9H1BM -122dBm

OM0WR -95dBm

DF2PY -88dBm

WA8OLN -78dBm

W3GH -60dBm

W4ZV -42dBm

The dynamic range between noise and W4ZV on one occasion was over 85dB! It also illustrates how important antennas, location, and propagation are rather than power. The dB difference between signals from the same area is profound. Many signals run at or near noise floor. This may not be typical of every night, but it shows how large the signal level variations between weak DX and strong signals are.

W8LRL and I have heard each other while running about 30 micowatts!

There are three main reasons we hear clicks and splatter off-frequency:

- The receiver is not up to the job, but it could be
- The transmitter is not up to the job, but it could be
- The signal is just too close in frequency, and reasonable use of technology won't cure the problem

Dynamic Range (IM3)

The most useful specification for receivers is called **dynamic range**. Dynamic range (DR) tells us the ratio in dB (anything described in decibels is a ratio) of the weakest signal that can be heard to the level where problems start.

Receivers

IM3 DR would be the ratio of two equal level signals (creating a third-order product by unwanted mixing) to the noise floor of the receiver.

Imagine we have two strong CW signals spaced 1kHz apart, one at 1840kHz and another at 1839 kHz. As the level is increased, the receiving system has a increasingly non-linear response. The 2nd harmonic of 1840 can mix with the fundamental of 1839, and the result would be a new signal at 2*1840-1839=1841kHz. Another signal, if the two original signals are equal strength, appears at 2*1839-1840=1838kHz.

(There are also sums, but they fall outside the filters and tuned circuits. We normally can't hear them, so we generally just ignore them.)

The response of this product is non-linear. The level of the mixing product increases faster than the level of either individual "real" signal.

When we can hear that "phantom signal" above the noise floor of the receiver, it adds interference

to other weak signals we might be trying to hear. We reference the main signal level this occurs at to the noise floor, because that is the level where it would start to be noticeable.

Always remember actual overload is the result of the vector sum of signals in the passband of the system at that point. One way to look at it is that overload is an accumulated power problem, not an absolute level problem with one signal. This means a great number of weaker signals can cause the same problem as a few strong ones.

Transmitters

Transmitters are a special case partially covered in a checking transmitters page.

Higher Order IM

We can have higher order IM products, but they are always odd sums or differences. For example the 2nd harmonic of one signal can mix with the third harmonic of another, and we have 5th order products (2*f1 minus 3*F2 called fifth because 2+3=5).

Although they can affect CW receivers, higher order products create most of our SSB problems. This occurs because higher-order products fall well outside the filter passband of a typical SSB receiver or transmitter.

Blocking

Blocking is the point where we can detect a change in sensitivity or gain from a single strong unwanted signal. Blocking "pumps" the receiver gain and can actually make false clicks. We always have to sure any click reports are given with signal levels well below the blocking point!

Blocking again is a ratio referenced to noise floor, since loss of sensitivity (either through increased noise or decreased gain) can affect readability of weak signals.

Transmitter Splatter

[Home]

Related pages: Checking signals Receivers Emission rules Part 97.307

Under construction 1/17/2004

Transmitter Intermodulation Distortion

A common culprit behind wide SSB signals (assuming the receiver is good) is high order distortion products. Transmitters process multiple frequencies at once. Mixing between these signals caused by amplitude response non-linearity can generate new frequencies. The offending products are those with odd-orders.

In a typical SSB transmitter the ***2nd RF harmonic*** of a modulating tone from one pitch can mix with the ***third harmonic*** of the RF caused by another modulating tone. This creates a new frequency that is called the ***5th-order product*** (2*F1 minus 3*F2 is called fifth because it is 2+3=5). We also can have a lower order product called the ***third-order product*** where the 2nd harmonic of one tone mixes with the fundamental of the other. We call that the third-order product because 2*F1 minus 1*F2 makes the new signal.

A 1850kHz LSB transmitter modulated with two tones of 500Hz and 3000Hz would have main signals at 1849.5 and 1847kHz. The third-order IM products would fall at:

(1849.5*2)-1847=1852kHz

(1847*2)-1849.5= 1844.5kHz

The fifth-order would fall at:

(1849.5*3)-(1847*2)=1854.5kHz

(1847*3)-(1849.5*2)=1842kHz

You can see every increase in order spreads the signal another (F1-F2) up and down the band. In

the above case F1-F2 is 2.5kHz. The 7th order product would be 2.5kHz above and below the 5th order products. Any odd product adds bandwidth to the signal.

If the product is direct mixing or even-order mixing, mixing would be (1*F1)-(1*F2), (3*F1)-(1*F2), (2*F1)+(2*F2), etc. The "harmonic order" in the mixing would total an even number. Let's try that:

(1849.5*1)-(1847*1)=2.5kHz outside the passband of the transmitter and antenna!

(1849.5*1)+(1847*1)=3696.5 outside TX passband.

(1849.5*3)-(1847*1)=3701.5 kHz again outside the passband of the antenna and transmitter's RF section. This was the fourth order.

Even-order isn't a problem in RF transmitters. This is why push-pull RF amplifiers don't help audible distortion and don't help bandwidth. They do cut down on harmonic distortion making band-filter requirements less.

Keep these rules in mind:

- ***Any increase in frequency difference between the highest and lowest modulation frequency increases BW greatly.***
- ***An increase in level increases the strength of the IM product in even greater proportion than we might expect.***

Higher odd-order products create most of our SSB problems. This occurs because high odd-order products fall well outside the normal passband of a typical SSB transmitter, spreading wasted energy to adjacent voice channels.

Three Greatest Sins

The three greatest sins creating unnecessary bandwidth are:

1. ***Turning up a radio's internal power or "drive limit" pot.***
2. ***Enhancing Bass and Treble at the same time***
3. ***Under-loading an amplifier***

Radios

Some modern radios are especially poor, even when operated at rated power levels. For example,

the TS-2000 and IC-756 series radios are particularly bad on 160 meters. I can hear some of these radios, when signals are strong, producing weak spurious emissions (splatter) 10-20kHz away from the operating frequency. If you look at ARRL test reports of transmitters, you will see many radios are just a bit above class-C performance levels. (Remember the ARRL uses dB below PEP, which improves results by 6dB compared to commercial test methods.)

IM3 levels of -30dB are really very poor. An old KWM2 I tested was -47dB using ARRL standards. Compared to something like an IC-756, the Collins had about ***50 times LESS power*** in total adjacent channel distortion products!

Compression and Processing

Any type of speech processing or compression (even ALC) makes the IM problem worse, because the average level of F1 and F2 increase. It is the multiple tones in our voices that create the "original" tones that mix, so the more we put in the more that come out!

Processing is a double-edged sword, however, because it also limits the level of peaks. Decreasing the level of peaks compared to average power loads bias and power supplies more steadily, and can keep stages after filtering out of limiting or distortion. While it might bring the level of "low-level nasties" up, it also decreases the likelihood of high-level overdrive problems.

Light or modest processing is actually beneficial in reducing splatter!

ALC

ALC is normally slow and plagued with inherent problems. Filters in radios add group delay (the signal takes noticeable time to move through filters), and the ALC loop adds a time-delay of its own. The result is leading edge signal overshoot, which often shows up as an adjacent channel "spit" or "pop" on leading or rising edges of voice or CW.

Some rigs like the early 775DSP and IC706 are plagued with very high levels of overshoot. A new 775DSP I had actually overshot to around 300 watts on leading edges for a few milliseconds. Kenwoods and other radios also have this problem. This problem is not only harmful for bandwidth, it also can damage amplifiers. The problem often gets *WORSE* as power level is turned down!

Gain should be set so ALC just starts to take effect, if a drive control is available. Rigs like the FT1000D include a "Drive" adjustment.

WideFi or Enhanced Audio

Enhanced audio is a generally bad idea, since it adds and boosts unnecessary lows and highs. The audio processing brings the level of the extreme low and high frequencies up to abnormal strength. Since the frequency spread of the lows and highs is wider and the level is stronger, IM products also are wider and stronger. Enhanced or Hi-fi audio, even with perfect "brick wall" filtering, is wider than communications audio for a given RF amplifier quality.

Many of the radios popular with the HiFi crowd are among poorer radios for IM performance! My own opinion is HIFi is OK on emptier bands, but we should do all we can to discourage enhanced audio on crowded bands or near weak signal areas.

Transmitter Tests

The standard transmitter test is IM3 or higher order products. The general test uses two tones of equal level. If it is a radio, the two tones are fed into the audio port. If it is an amplifier, the tones must generally be from two separate transmitters generating steady carriers. The carriers are mixed through a combiner and used to drive the amplifier. The reason we use two separate transmitters in the test is most amplifiers, especially cathode driven triodes, are far cleaner than most modern radios. The test radio's two-tone IM would establish the IM distortion limit (not the amplifier, in most cases.

Confusing Conflicting Standards

In most commercial tests, we find the maximum power of one tone and compare that level to the level of third-order spurious signals created in the transmitter. The ARRL, for some reason, adopted a different reference. They compare the PEAK power of both tones to the spurious, rather than the level of one test tone. This inflates transmitter IM results, making everything appear 6dB better than it really is in the standard test used commercially.

If you look at Eimac data sheets, the IM specs are dB-below-one-tone of the two tones. If you look at data sheets from other sources, they might use dB below PEP. One manufacturer's US importer of Russian tetrodes used dB below PEP to compare the quality of their product to Eimac.

Unfortunately Eimac used dB below one tone, while the other tetrode test used dB below one tone, so the tetrode manufacturer wrongly proclaimed their tetrodes "cleaner" when in fact they were not. If you look at QST tests of ETO, QRO, and Ameritron amps you will see the 3CX800's in the AL800H greatly surpass the 4CX800's in the other amps for IM3 and IM5. This confusion is a clear example of how mixed standards gets us in trouble.

It's a Poor Test Anyway

Two-tone IM3 (or higher orders, like IM5) transmitter tests generally show us the very ***BEST*** a

transmitter will do. In general, two-tone tests are pretty poor tests for system designed to process speech. The two-tone test uses two steady signals (normally with wide spacing), but actual modulation has many frequencies that vary at syllabic rates.

In two-tone tests, the test spacing is generally a few kHz. The varying load on power and bias supplies is at the separation of the two frequencies. Small capacitors filter the time-varying load, while the long term dynamic load remains constant. Two-tone tests do ***NOT*** show power supply deficiencies.

Slow variations in speech level load and unload power and bias supplies. This causes supplies to "wobble around". The conclusion of some is that screen or bias regulation is "unimportant", but that conclusion is mostly rooted in the fact they use a very poor test method that does not show low frequency dynamic regulation problems.

A Better Test

There are two tests that are better. One test is an adjacent channel power tests, with normal voice modulation of the transmitter, another test I developed uses a three-tone signal.

The three tone test injects a third low-frequency tone into the system. Is anything from a warble to a low-pitched hum variation in power levels of the two major tones. The analyzer reads the peak amplitudes of mixing in higher frequency tones while the level is varied at a syllabic to low pitched audio rate. The low pitch amplitude modulation is varied in frequency until the worse case IM is produced.

The variation causes the power to change at a speech rate, testing supply regulation effects on wide spaced distortion at all important frequencies for speech.

The adjacent channel power test simply uses normal voice operation and compares the long term peak power in an adjacent channel to the peak power in the desired channel.

Either test gives a much more reliable indication of transmitter bandwidth than a two-tone test.

Remember, a two-tone test is generally a "best case" scenario.

Summary

A normal two-tone test is not really very effective in measuring a SSB signal, because there is no slow dynamic change typical of a voice. Voltage regulation problems are masked and do not show up when a two-tone test is used, because load currents all average the same amount. Filter capacitors and other energy storage components mask any voltage regulation problems. We can, as

a general rule, be confident the actual SSB voice performance is LESS than a two-tone test indicates.

A large improvement occurs with a three-tone test, when levels are varied at syllabic rates as well as higher frequencies in the voice range. This tests voltage regulation problems otherwise hid in a two-tone test, because all speech frequencies ranges are included. The most accurate test, of course, is with actual speech. The FCC now requires some commercial radios used on congested bands to be tested with actual speech.

We can help ease spectrum pollution by:

- **Strongly discouraging use of enhanced bass and treble on crowded bands. It has no place on crowded band or near weak signals. Wide-Fi is selfish and inconsiderate even when it uses a 3kHz filter because of the increase in level and frequency spread of IM products.**

- **Discouraging and chastising people would turn the power limit or drive limit control inside radios up. This is CB behavior! There isn't a radio made that will tolerate a user-increase in power limit without a serious degradation in IM performance. If you have a friend who peaks up the power control inside a radio, tell him why it is bad. Radios are bad enough without removing even more headroom. Even the best transistors cannot be driven more than about half of saturated power before IM becomes unacceptable.**

- **Making sure we use processing and ALC, but only at modest levels. We should just see the needles start to show compression.**

- **Making sure we tune amplifiers correctly, and avoiding 12 volt transistor amplifiers or grid-driven tetrode amplifiers whenever possible.**

MFJ-259B Calibration

[Home]

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Revised 8/24/03

History:

I worked with another contract engineer (my friend JB) in designing the MFJ 259 and 269. No one else outside of MFJ was involved, and certainly no one named Ted Hart.

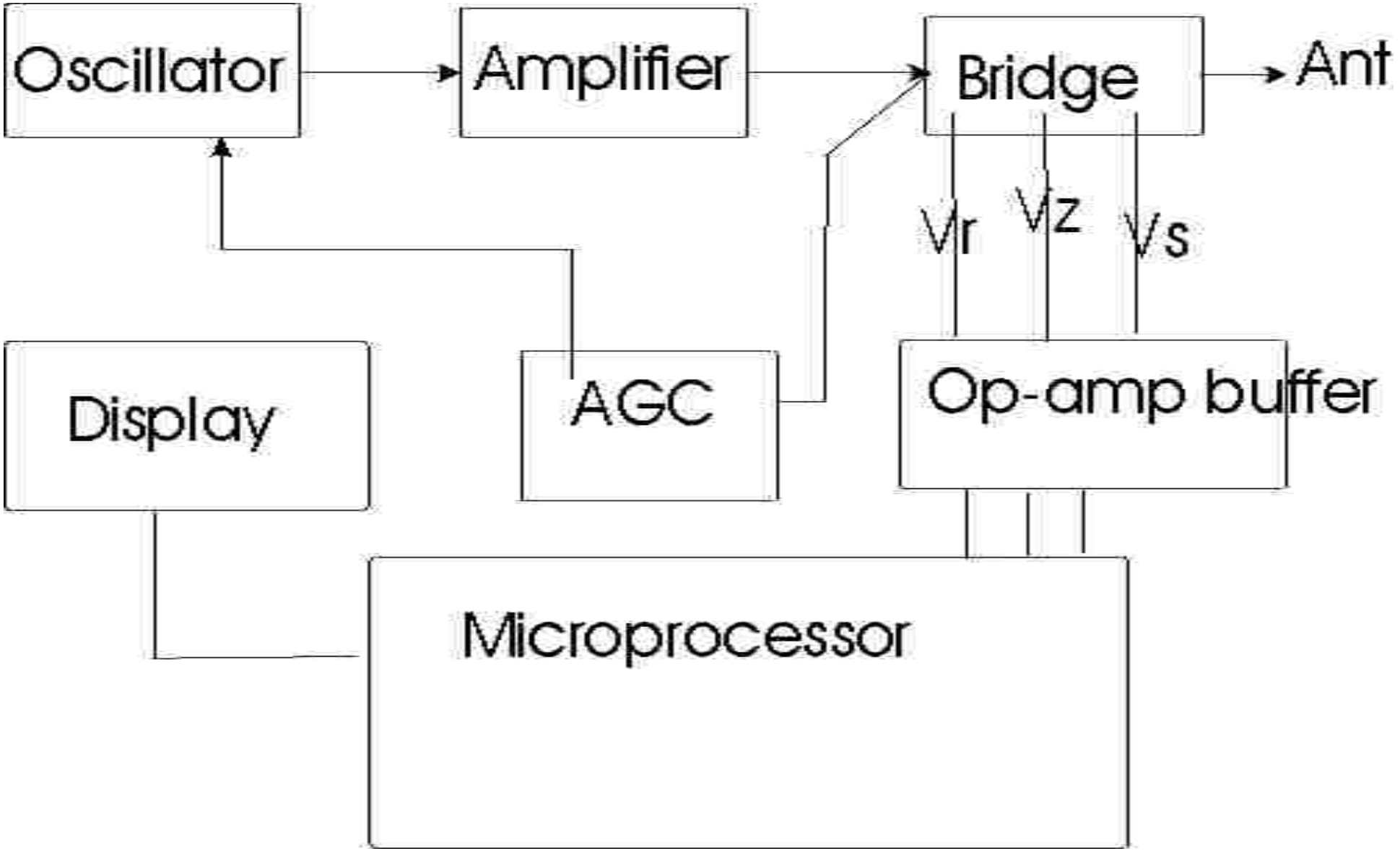
This information is here because it is the correct way to calibrate the MFJ-259B analyzer. This work is all donated. K1BQT took an instruction set supplied by MFJ and re-wrote it. I modified, expanded and edited that work. This page is the result.

It is best that no one copy this, and start handing it out in mass. The only reason for this request is there must be a point of control of information, so it can be corrected or expanded as errors or omissions are found.

I am not aware of any other source that gives correct calibration procedures. It is important that the 259 be calibrated by these steps, even if they sound complex. If you don't do it right, don't do it. Without following these steps, many special functions may not work correctly even if the unit tests properly on calibration loads!

How This Type of Device Works

This type of analyzer contains an RF oscillator, a linear amplifier to increase power, and an internal resistor bridge in a conventional Whetstone bridge configuration.



Since it is designed to be inexpensive, there are a few shortfalls with this system.

The bridge is dc-coupled from an internal resistor bridge to the antenna port. The bridge detectors are NOT frequency selective, and respond to anything from minor dc offsets through microwave signals. This causes a potential problem if there is ***any*** voltage appearing across the antenna port, from dc through microwave. (This is also true for competing analyzers from other manufacturers.) There are multiple reasons why, at the time of design, these units were dc coupled with broadband detectors. Hopefully someday a higher cost-design with selective detectors will become available, but for right now

this is all that is available for amateur use from any manufacturer.

The second shortfall is the internal amplifier must be linear and have very low total harmonic content. Total harmonic power, at the lowest load impedance, must be down at least 25dB and preferably 35dB. This is true for ***ANY*** antenna analyzer, since you do not want the analyzer to measure the load at two frequencies!

Because the detector is broadband and because it is dc coupled to the antenna, any external voltage across the input port causes measurement errors. It is the **accumulated voltage of multiple sources** that is most important, not the strength of any individual signal. Because of that, large antennas should be tested at times when propagated signals in the range of the antenna's response are at minimum strength.

A definite RFI improvement occurs with a bandpass filter, but multiple-section bandpass filters cause impedance measurement problems. Multiple-section filters behave like transmission lines of random line impedances, loss, and lengths as frequency is varied. The best solution is to use a single-stage bandpass filter and dc isolation on large arrays or with long feedlines. I often use a good 1:1 isolation transformer for measurements, and often find a parallel L/C filter (like the MFJ-731 Filter) useful.

Most Likely Failures

Other than manufacturing errors, the **detector diodes clearly stand out as the most common problem. They are the most easily damaged devices in the analyzer.** If you have a sudden problem, it is most likely a defective detector diode. Diode damage almost always comes from accidentally applying voltage on the antenna port.

Why are the diodes so sensitive? In order for the detectors to be accurate within a fraction of a percent (one bit), detector diodes must have very low capacitance and a very low threshold voltage. This means the diodes, through necessity, must be low-power zero-bias Schottky microwave detector diodes. The same characteristics that make them accurate and linear cause the diodes to be especially sensitive to damage from small voltage spikes. ***ALWAYS discharge large antennas before connecting them to the analyzer! Never apply external voltages greater than 3 volts to the antenna port!***

Technical Support

From time to time MFJ gives incorrect advice on the 259B and other analyzers. One bad piece of advice that has come to my attention concerns measuring stubs. Someone in MFJ support has been telling customers that the manual is wrong, but the manual is actually correct! The person in support is wrong.

A problem with measuring stubs and cable lengths occurs if the harmonics are high in the 259 or 269, or if you do not tune for lowest X and minimum impedance and store that impedance, and then move to the VERY NEXT minimum X and lowest impedance. Whatever anyone tells you at MFJ, that is how it has to be done.

How This Unit Works

This is a rough outline of how this unit works:

The MFJ-259B, and other digitized MFJ antenna analyzers, compare three major voltages in a 50-ohm bridge circuit. They are:

Vz= Voltage across the load. This voltage is called Z in the alignment display menu

Vr= Voltage indicating bridge balance. This voltage is called R in the alignment display menu

Vs= Voltage across a 50-ohm resistor between the RF source and the load. This voltage is called S in the alignment display menu

All voltages are converted through an eight-bit A-D converter to a 256-bit digitized output with a test-display range of 0-255 bits. By knowing the ratio of these voltages, as compared to the regulated RF source voltage, many different load parameters can be calculated. An antenna analyzer could calculate everything (except sign of reactance) from measuring only Vs and Vz, but at certain impedances any small error in either Vs and Vz becomes critical. This is especially true when voltage is digitized into a 256-bit format (~0.4% steps). At certain impedances, an almost immeasurable voltage change will cause a sudden large jump in the measured impedance parameters.

To reduce display impedance jumps, SWR is weighed into the calculation of reactance and resistance at low SWR values. (An SWR bridge is most accurate when the load is closest to 50 ohms, which is a primary measurement area where impedance measurements through Vz and Vs become critical.) By factoring in a direct SWR measurement from an internal bridge, the analyzer can check and "correct" any small level errors in Vs or Vz. This reduces the impedance jump that would occur with a one-bit jump in voltage. This also why bits must be calibrated for near-perfect accuracy. a one-bit error can cause a resistive load to appear reactive (the total of Vs and Vz must always be 255 bits or less for a load to be considered resistive).

Calibrating the MFJ-259B Antenna Analyzer

This calibration procedure is the correct procedure for later MFJ-259B's. Disregard any other information. Since MFJ-259B firmware has several versions under the same model number, you may find some final performance verification steps not valid. These steps will involve parameters that do not appear on the display.

Be sure you have printed a copy of the board layout showing adjustment points, have read all this, and have suitable loads before proceeding.

Adjustments

This unit has tracking and gain adjustments for Vz, Vs, and Vr. Tracking is set at low voltages (low bits). Gain is set at high voltages or bits. Together they make the detector voltage output closely track the actual RF voltage.

This unit also has meter calibration adjustments. The analog meters almost certainly suffer from some scale linearity problems, they will be somewhat less accurate than the digital display. These adjustments only affect analog meter readings. The meter adjustments do not affect the display.

Quiescent current (bias) in the RF amplifier section is adjustable. This adjustment directly affects output signal harmonic content. Harmonics are worse with low supply voltages, and with low impedance loads. ***Be sure you check the harmonics as outlined below, with a 1/4 wl open-circuit stub!!***

Excessive harmonics can cause severe errors in measurement of frequency-selective loads, even when dummy-load SWR tests appear perfect. Loads most sensitive to harmonic-induced errors include, but are not limited to, antenna tuners, tank circuits, very short resonant antennas, and distance to fault and stub length measurements. If you notice something "funny" going on with a stub measurement, it may be a fault of incorrect bias.

Warning: Never calibrate around a sudden "problem" that appears. If a detector suddenly shifts voltage, the problem is almost certainly a defective detector diode. If the meter is recalibrated with a defective (leaky) diode, the meter will probably NOT track correctly with frequency.

Alignment

Tools and Equipment:

[ ] #2 and #1 Phillips-head screwdrivers

[ ] Digital meter or accurate analog meter for checking supply voltage

[ ] Small set of non-metallic alignment wands for coils, or small jeweler's screwdrivers for controls

[ ] Power supply, regulated to 12-volts + - 5%

[ ] General-coverage receiver with level meter or Spectrum Analyzer (these are optional with additional work and use of a stub)

[ ] ~10 MHz 1/4wl open-stub, for example 15’ good-quality solid-dielectric RG-8, UHF connector at on end, open on other (not needed with analyzer or receiver)

[ ] 2.2-ohm 1/4 or 1/2 watt film resistor (not needed with stub)

[ ] Accurate load set to include:

A. Short

B. 12.5-Ω load

C. 50-Ω load

D. 75-Ω load

E. 100-Ω load

F. 200-Ω load

Note 1: Loads must be constructed using physically small 1% carbon-film resistors.

- **DO NOT use large resistors. Acceptable results will be obtained when load resistors are mounted in the very bottom of a UHF-male connector.**
- **The ideal load resistors are surface-mount precision-resistors, but other styles will work. It is acceptable to parallel multiple resistors to obtain low resistances, but don't series connect more than two resistors!**
- **Never use physically large resistors, such as 1-watt or larger resistors, unless you are absolutely positive they are composition-types (very rare).**
- **Since the loads are used to set the number of bits in critical calculations, the maximum reactance error will always be worse than the percentage of resistive load error. *A one bit error in calibration (~ .4%) can cause a purely resistive load to read reactive.***

Quick-connect loads can be made with surface mount resistors on a BNC male chassis mount connector with the bayonet removed. This makes a "quick connect" connector that will slide directly into a type-N female. In this case, use a good UHF to BNC female adaptor for the 259 units. With a 269, the load will plug directly in to the N female.

Note 2: The power source should be the LOWEST expected operating voltage. DO NOT use a standard "wall-wart" or batteries! You can reduce voltage from a conventional 13.8v regulated supply by adding a few series diodes. Silicon diodes will normally drop about 0.6volts or so per diode. Three or four diodes will reduce place the voltage below 12 volts.

WARNING: The MFJ-1315 AC adapter or other "wall-warts" should NOT be used to power the unit for most alignment steps.

Step 1

Visual Inspection: Before, during, and after calibration, be mindful of physical condition. Watch for missing or loose hardware. Do not tug, stress, or repeatedly flex leads, or carelessly flop or toss things about. Keep your bench clean. Follow these rules the entire time you have the unit apart!

Step 2

Battery Tray Removal: This step provides access to trim-pots and most inductor adjustments.

[ ] Remove last two batteries at each end of the tray.

[ ] Remove two screws (right side) and extract the tray.

[ ] Always position the battery tray to minimize strain on wires.

Refer to the board layout pictorial for specific control locations.

Step 3

Band Overlapping: Each band should overlap the next by a small amount to ensure gap-free coverage from 1.8 MHz to 170 MHz. While viewing the *LCD Frequency Display*, wiggle the bandswitch from side-to-side gently. Watch for any display or meter dropout. Check each band as follows:

[ ] **114-170 MHz:** Oscillator tunes from below 114.0 MHz to above 170.0 MHz. Check tune for dead spots.

[ ] **70-114 MHz:** Oscillator tunes from below 70.0 MHz to above 114.0 MHz

[ ] **27-70 MHz:** Oscillator tunes from below 27.0 MHz to above 70.0 MHz.

[ ] **10-27 MHz:** Oscillator tunes from below 10.0 MHz to above 27.0 MHz.

[ ] **4-10 MHz:** Oscillator tunes from below 4.0 MHz to above 10.0 MHz.

[ ] **1.8-4 MHz:** Oscillator tunes from below 1.8 MHz to above 4.0 MHz. Check tune for dead spots.

*While verifying overlap, at least check the lowest and highest bands carefully for dead spots. The LCD Display will indicate 000.000MHz if a dead spot occurs. Dead spots generally indicate a defective tuning capacitor (**TUNE**).*

If you find wiggling the switch causes a dropout, the switch may have dry or dirty contacts. Less likely are poor solder joints, but check solder joints first. If you must clean and lubricate the switch, be aware it is a difficult task. The entire board needs to be lifted from the case front. Dirty band-switch contacts may be restored with spray tuner cleaners. The best place to spray the switch is from the front side (shaft side), right below the nut. You must remove the switch indexing tab retainer nut and the metal switch retainer (stop) under the nut. Be sure the stop goes back exactly as removed.

To correct overlap problems, locate and retune the appropriate VFO coil (see Pictorial for coil locations). Note that L1-L4 are slug-tuned and require an insulated hex-head tuning wand. If you use the wrong size wand or a worn wand, it will break a slug!

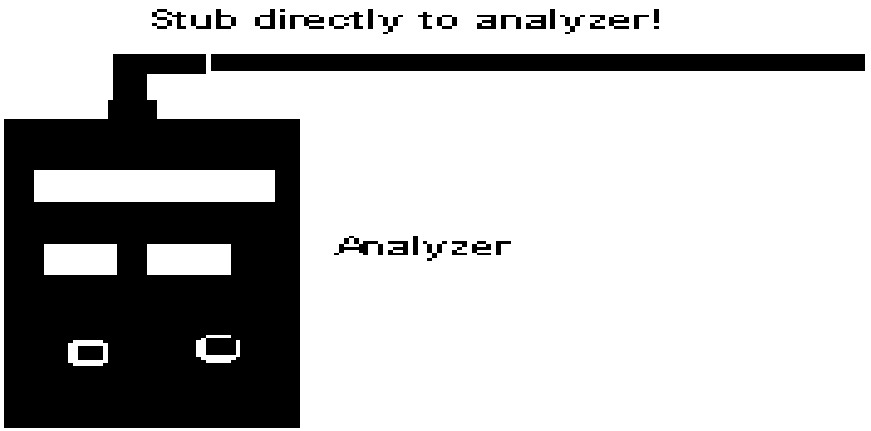
Inductors L5 and L6 are located on the component side of the board and are compression-tuned (press turns closer together to lower frequency or spread apart to raise frequency). Make only very small corrections--*especially to L5 or L6*--and recheck the band you are adjusting. You should also check the *next lower band* after each adjustment to ensure that the lower band hasn't moved excessively.

Important Warning: VFO coils MUST be aligned from highest frequency to lowest frequency. The next higher range affects next lower band the greatest amount. Do not attempt VFO coil adjustment unless you are experienced working with VHF-LC circuitry or complex alignment procedures.

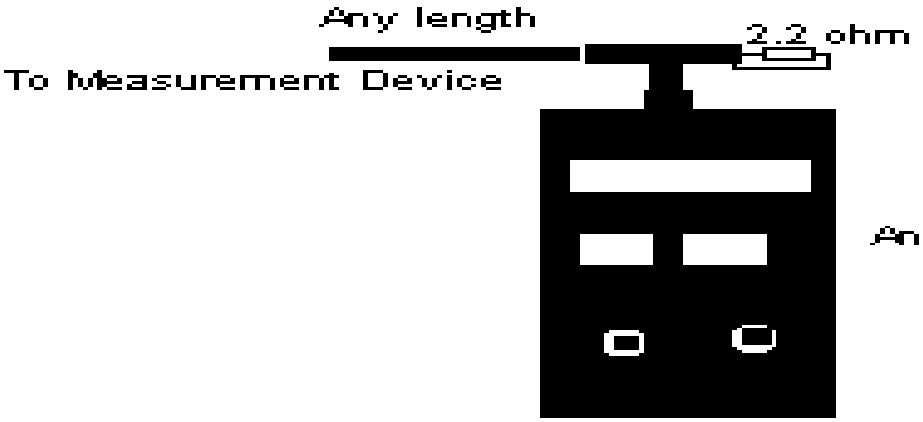
Step 4

Harmonic Suppression/ Bias: Connect the analyzer ***exactly*** as shown below.

- The impedance of the cable to the measurement device should match the impedance of the measurement device.
- The "T" must be connected either directly to or placed within a few inches of the analyzer.
- The power source must be the lowest expected operating voltage.
- The measurement device must be well-shielded, and not pick up any substantial signal from the analyzer when the "T" is disconnected from the analyzer.



Test Setup Using Internal V2 for Harmonics



Test Setup using External Analyzer and resistor for Harmonics

Step 5

Harmonic Suppression (bias R89): This adjustment reduces oscillator harmonics. Harmonics will cause incorrect readings under some load conditions.

WARNING: ***Incorrect adjustment of R89 will NOT show when checking with resistive dummy loads!!! The unit will appear to calibrate correctly, but will produce errors in stub length, distance-to-fault, and other frequency selective functions.***

When R89 is set properly, harmonic suppression of –30 to –35dBc should be possible across most of the analyzer's tuning range. This particular

adjustment must be made at the lowest expected operating voltage. Proper alignment requires a 12.0-volt regulated supply as a power source. NEVER use an AC adapter or any supply voltage higher than 12-volts when making this adjustment. A calibrated spectrum analyzer works best for monitoring harmonic output, but a well-shielded general-coverage receiver with signal-level meter will also work. The receiver MUST be "T'd" into the analyzer just as the spectrum analyzer is, and the Tee and resistor must be located at the analyzer connector. If you do not have a good-quality receiver or spectrum analyzer, use the test mode of the analyzer with a stub. Watch MFJ analyzer test-mode Vz. Test-mode Vz will roughly indicate total harmonic voltage when the analyzer is set at the stub's exact resonant frequency. Entering the test mode is described in Detector Calibration (Step 6).

[ ] **a.** Install either a 15' RG-8 open stub, or resistor and measurement device, and tune analyzer to approximately **10 MHz**.

[ ] **b. (stub and internal Vz use only)** Observing Vz on the data display (analyzer test mode), adjust frequency until the *lowest* fundamental output reading (or lowest impedance) is obtained. You should clearly see the MFJ analyzer's fundamental frequency output voltage (Vz) go through a deep null.

[ ] **c.** Observe the analyzer frequency reading. This is the approximate resonant frequency of the stub, and the test frequency.

[ ] **d.** Without changing the analyzer test frequency setting, observe the second harmonic level. This harmonic will be at twice the MFJ analyzer frequency counter reading..

[ ] **e.** Adjust **R89** for **lowest 2nd harmonic meter reading** on the receiver, lowest Vz test-menu reading, or lowest harmonic level on the spectrum analyzer. Be SURE the fundamental frequency level remains nulled in the analyzer.

WARNING: Always repeat steps (b) through (e) at least one extra time when relying on display Vz. The original null point of any stub will shift if there is a substantial reduction in harmonics after R89 is adjusted. The original stub frequency, as observed at (c), will probably change slightly. It is NOT necessary to recheck when doing a resistor load test with a good-quality spectrum analyzer or receiver. With a resistor, exact test frequency is NOT critical.

NOTE: If you have a poorly performing spectrum analyzer or receiver with limited dynamic range, use a stub with the spectrum analyzer or receiver instead of a 2.2 ohm resistor. If you have a reasonable quality spectrum analyzer or receiver (at least 50dB dynamic range) use a 2.2-ohm non-inductive resistor in lieu of the stub, resistor adjustment is easier and more accurate.

Detector Calibration

Step 6:

This critical sequence calibrates A-D conversion for various load conditions. If you know your unit has been tampered with, preset trim pots **R88**, **R89**, and **R90** to their center positions before continuing. If you find any control bottoms-out in adjustment, you almost certainly have installed an incorrect load or the analyzer has a defective detector diode.

To prepare for detector tracking alignment, place the analyzer in *Test Mode*. Entering test mode may be tricky with some units, and it may take practice. To enter *Test Mode*:

[ ] Turn power off.

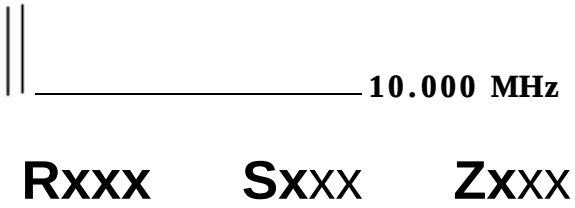
[ ] Hold down ***MODE*** and ***GATE*** buttons while restoring power.

[ ] As display comes up, slowly (about 1 second period) rock between pushing the **MODE** and **GATE** switches alternately (the best method is to use two fingers, and rock your hand from side to side between the two buttons)

[ ] Confirm analyzer has entered test mode (it may take more than one try).

[ ] Using the **MODE** button, advance display to the **R-S-Z** screen (shown below).

Note: If you go past the R-S-Z screen, you can still see R-S-Z by pushing and holding the MODE button.



[ ] Tune analyzer operating frequency to approximately **10.000 MHz**

[ ] Leave antenna connector **Open**

[ ] Set **R72** for **Z=255**

[ ] Install the **Short**

[ ] Set **R73** for **S=255**

[ ] Install **12.5-Ω** load

[ ] Set **R90** for **Z=051**

[ ] Set **R53** for **R=153**

[ ] Install **200-Ω** load

[ ] Set **R88** for **S=051**

[ ] Set **R72** for **Z=204**

[ ] Install **75-Ω** load

[ ] Set **R89** for **R=051**

[ ] Install **12.5-Ω** load

[ ] Reset **R90** for **Z=051**

[ ] Set **R73** for **S=204**

[ ] Reset **R53** for **R=153**

[ ] Install **200-Ω** load

[ ] Reset **R88** for **S=051**

[ ] Verify or set **Z=204**

[ ] Install **75-Ω** load

[ ] Reset **R89** for **R=051**

Important Note: **Small single-turn trimpots are touchy to adjust and tracking settings are somewhat interactive. If specified readings aren't obtained on the run-through, repeat the sequence a second time (accuracy counts). When the sequence is complete, turn power off to remove the analyzer from *Test Mode*.**

Be particularly mindful of the total bits of Vz and Vs. If the sum of these bits ever exceeds 255 with a resistive load, the analyzer

will indicate reactance.

Display Test and Analog Meter Calibration

Step 7:

This step sequence checks meter calibration and verifies accuracy of the *LCD Display* information.

Remove and re-apply power and enter the *Real-Imaginary* impedance mode **R-X**. Readings + or - 10% of reading or + or - 5 ohms of display, whichever is larger, are considered within design specifications. Typically digital readings are almost perfect with proper detector calibration. Analog meter readings may be outside that range, and as much as 20% off with some load values.

[ ] Install **75-Ω** load

[ ] Verify reading of **R= 75 X=0** on *LCD Display* (±10%)

[ ] Install **50-Ω** load

[ ] Verify reading of **R=50 X=0** on *LCD Display* (±10%)

[ ] Set **R67** for reading of **50** on the *Impedance Meter*.

[ ] Verify reading of **1.0** *SWR Meter* (no deflection).

[ ] Install **Open** load

[ ] Verify reading above **400** on *Impedance Meter*

[ ] Install **100-Ω** load

[ ] Verify reading of **R=100 X=0** on *LCD Display* (±10%)

[ ] Verify reading of **100** on *Impedance meter* (approximate).

[ ] Set **R56** for a reading of **2** (2:1) on the *SWR Meter*

[ ] Install **12.5-Ω** load

[ ] Verify a reading of **4:1** SWR on *LCD display* (3.8-4.2 good)

[ ] Verify reading of **>3** (greater than 3:1) on *SWR Meter*

[ ] Install **200-Ω** load

[ ] Verify reading of **4:1** SWR on *LCD display* (3.8-4.2 good)

[ ] Verify reading of **>3** (greater than 3:1) on *SWR Meter*

Capacitance Mode Check

Step 8:

If you have a few precision capacitors, you can verify the calibration between the ranges of 100 and 5000 pF. Read the analyzer manual for details of capacitor measurement.

[ ] Install no load

[ ] Set *Mode* to *Capacitance*

[ ] Set VFO to **70 MHz**

[ ] Verify **4-6 pF** reading on LCD Display

Frequency Counter Check

Step 9:

These steps verify accuracy of the counter. Note that the counter's clock isn't user-accessible, so no adjustments will be made. To conduct this test, use a general-coverage receiver in AM mode.

[ ] Tune in **WWV** on 5.0,10.0,15.0, or 20.0 MHz (frequency with best reception).

[ ] Install a short clip lead or wire in the analyzer's *Antenna* jack.

[ ] Turn on the analyzer and **zero-beat the WWV signal** as closely as possible.

[ ] Compare *LCD Display* reading to the WWV frequency being used.

[ ] Verify agreement is within ±**5 kHz.**

Advanced Modes Check

Step 10:

This sequence verifies operation of the analyzer's advanced features. To enter *Advanced Mode* menu:

[ ] Turn unit off.

[ ] Hold down the MODE and GATE switches while turning power on.

[ ] Verify "*Advanced"* appears on the *LCD Display*.

[ ] Install **Open** load

[ ] Tune VFO to **>170 MHz** (top end of coverage range)

[ ] Verify **Z** = **<650 Ω** with (about) **90**° phase shift

[ ] Install RG-8 open stub

[ ] Tune VFO for minimum **Z** reading (around 10 MHz)

[ ] Verify **Z-min = 0** to **2 Ω**

[ ] Install **50-Ω** load

[ ] Set VFO to **1.8 MHz**

[ ] Verify **Z = 50** Ω, θ = **0**°, and **SWR = 1** (±10%)

[ ] Enter **RL** *Mode* (return loss)

[ ] Verify **RL = >42 dB**, ρ = **0**, **SWR = 1**

[ ] Enter *Match Efficiency Mode* (skipping *DTF Mode*)

[ ] Verify **ME ≅ 100%** (approximate)

[ ] Press and hold MODE and GATE buttons to restore *Main Modes*

[ ] Remove load and verify **Z = >650** on *LCD Display*

Conclusion:

Step 11:

[ ] Reinstall battery tray

[ ] Confirm charger jumper is set for type of batteries used (disable for alkaline).

[ ] Reinstall cover

This completes calibration.

MFJ-259B Calibration Checklist

Make a copy and check each box as you proceed down the calibration list.

Physical Condition

[ ] Hardware, batteries okay

Harmonic Check

[ ] Suppression -35 dBc or better

Overlapping

[ ] All bands have sufficient overlap

Check Digital Calibration at 10 MHz in test mode.

[ ] **Open**

[ ] R72 for Z=255

[ ] **Short**

[ ] R73 for S=255

[ ] **12.5-Ω**

[ ] R90 for Z=051

[ ] R53 for R=153

[ ] **200-Ω**

[ ] R88 for S=051

[ ] R72 for Z=204

[ ] **75-Ω**

[ ] R89 for R=051

[ ] **12.5-Ω**

[ ] R90 for Z=051

[ ] R73 for S=204

[ ] R53 for R=153

[ ] **200-Ω**

[ ] R88 for S=051

[ ] Verify Z=204

[ ] **75-Ω**

[ ] R89 for R=051

Analog Calibration: 10 MHz, values approximate

[ ] 75-Ω

[ ] Verify R= 75 X=0

[ ] **50-Ω**

[ ] Verify R=50 X=0

[ ] Set R67 for 50 on Imp Meter.

[ ] Verify 1.0 on SWR Meter

[ ] **Open**

[ ] Verify >400 on Imp Meter

[ ] **100-Ω**

[ ] Verify **R=100 X=0**

[ ] Verify 100 on Imp Meter

[ ] R56 for 2 (2:1) on SWR Meter

[ ] **12.5-Ω**

[ ] Verify 4:1 on LCD (3.8-4.2)

[ ] Verify >3 on SWR Meter

[ ] **200-Ω**

[ ] Verify 4:1 on LCD (3.8-4.2)

[ ] Verify >3 on SWR Meter

Capacitance Mode Check

[ ] **Open**

[ ] Set VFO to 70 MHz

[ ] Set Mode to *Capacitance*

[ ] Verify C≅4-pF

Counter Check

[ ] Counter Okay

Advanced Modes

[ ] Tune to 170 MHz

[ ] **Open**

[ ] Verify <650, Phase ≅ 90°

[ ] **3' RG-58**

[ ] Tune for Z-min (≅150 MHz)

[ ] Verify Z= 0-2 Ω

[ ] **50-Ω**

[ ] Tune to 1.8 MHz

[ ] Verify Z=50 Ω, θ=0°, SWR=1

[ ] Advance to Return Loss

[ ] Verify RL=>42dB, ρ=0, SWR=1

[ ] Advance to Match Efficiency

[ ] Verify ME ≅ 100%

[ ] Restore Main Modes

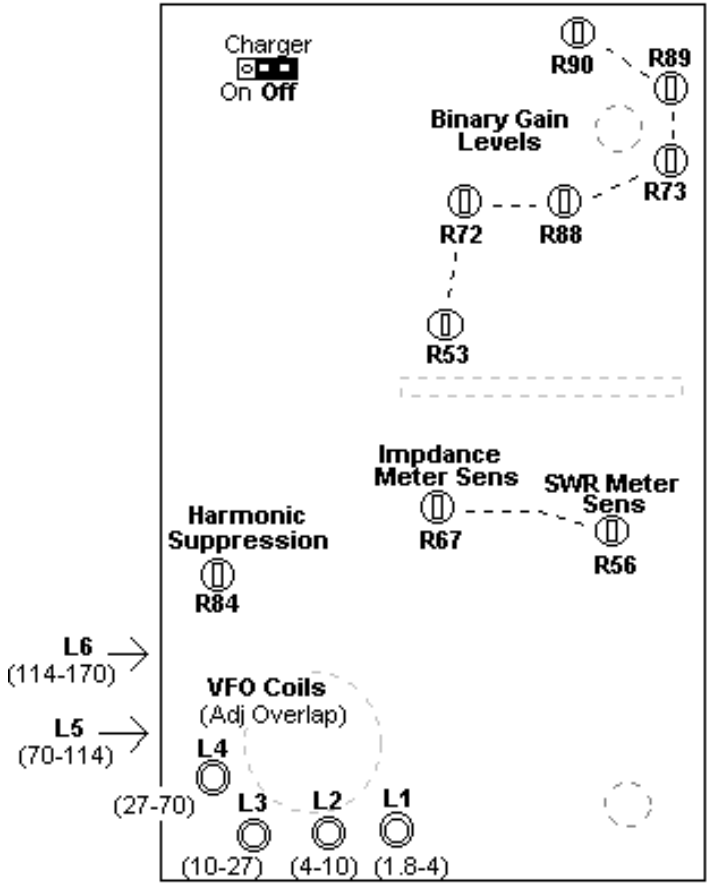
[ ] **Open**

[ ] Verify Z=>650

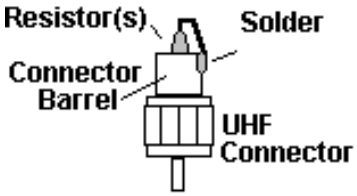
End of Procedure

Pictorial Diagram of Analyzer Board

Locations for trimpots and inductors



Loads Using Standard-Value Resistors



- Install resistors all the way down in the connector, the goal is zero lead length

.

- Use precision 1% carbon or metal film 1/8th-1/4 watt resistors

12.5Ω = (4) 50-ohm or a single 15Ω and 82Ω 1% in parallel

50Ω = 49.9-ohm or 100Ω and 100Ω in parallel

75Ω = 75-ohm or 150Ω and 150Ω in parallel

100Ω = 100Ω

200Ω = 200-ohm or 100Ω + 100Ω in series

Important Note: *These simple HF loads will not always be accurate for SWR checks above 30 MHz. Only precision terminations should be used in the VHF region, and even then there can be some errors. The MFJ-259B does not correct for connector impedance bumps or the electrical length between an external load and the bridge inside the unit.*

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RF Noise Powerstroke Diesel

[Home]

Related pages:

mobile antennas Loading coils Field Strength Measurements

RFI measurements showed HF noise at a distance of 5 meters from my 2003 Power Stroke was 15-20dB less than my 1995 F-250. I never could get all the noise out of my 7.3L 1995 F-250 HD PSD. The measurements of noise field a small distance from the truck gave me great hope I could have a better HF mobile with minimal work.

I am delighted to report my 2003 F-250HD PowerStroke diesel has virtually no RFI with only very simple and fast corrections. Be aware this is for my antenna mounting location, and my style of truck. This is a new old-stock truck I found at a small rural dealer for a *very* good price!

Antenna Mounting Location

I mounted the antenna in the stake hole of the bed that is located just behind the cab on the driver's side. I use this location because:

1. I do not want to drill holes in a new truck
2. I have no topper or camper shell to mount the antenna to
3. I want maximum performance
4. I would like to remove the mount if necessary

Antenna Mount

My antenna mount is home made. It is formed from a scrap sheet of stainless steel that looks to be about 7 gauge. In order to make the mount, I went though the following process:

1. bent two angle-sticks of metal that match the rectangular dimensions of the stake hole
2. TIG welded a 3/4inch nut inside one L to match the height of the existing Ford bolt hole
3. TIG welded the L's into a rectangular tube
4. TIG welded a few sharp points on the tube near the bolt hole to break the paint for a ground connection
5. bent a wide long plate into an L that matched the size of a Tarheel mount
6. added holes to match the Tarheel plate
7. TIG welded the bent plate to the end of the tube
8. added gussets to brace the bend, since that was a weak area that allowed the antenna to flop around too much
9. cut a 4" square load spreader out of 3/16 inch thick scrap stainless
10. drilled a hole in the load spreader plate
11. cut square of rubber as a buffer to prevent damage to the trick's bed rail top

You can see where I get my shield ground connection. I feed the antenna with the single wire to the lug normally used to mount the impedance correction coil. I mount my impedance correction coils on PL-259 plugs so I can change them easily. I screw them into the SO-239 connector normally used as a feed!

Noise Noise Noise

When the antenna was installed, I had S-7 injector noise on my IC-706. I cured it with the following:

I removed the Tarheel motor line choke bead because it was far too low an impedance. I was receiving on that lead, and it was bringing RF back into the cab, I use my own jumbo size 44 mix bead.

You will probably find most frame-type vehicles mount passenger compartments on rubber mounts to reduce noise and vibration. The cab is only grounded at the front. The bed is bolted solidly to the frame, and this means all RF currents must travel down the bed to the frame, follow the frame forward to the front of the cab, flow back along the cab towards the antenna. This actually makes a very short thick antenna out of the truck frame and cab of the truck! The common mode currents excite all the wiring under the vehicle, increase ground losses under the truck, and increase noise ingress from truck wiring into the antenna.

To cure the common mode current problems, I added a ground strap between the bed and the cab.

You can make the connection UNDER the chassis at the locations shown in the following picture:

Note that I heavily tin the edges of the 1" braid. I do this insure all of the braided wires stay fully in contact all of the time. The flex area must remain solder free.

I use stainless thread cutting screws, and stainless external tooth star washers under the braid in the tinned area to cut through the paint.

Adding this single strap made the following changes:

The base resistance of the antenna dropped from 45 ohms to 25 ohms on 7 MHz. This mean I greatly reduced radiation from the truck body or frame.

Noise level dropped 2-3 S units

RF in truck wiring dropped substantially

Remaining Noise

A small amount of noise remained audible after the grounding. I traced this noise to radiation from the exhaust system. The engine block was

hot with RF, the injector control modules are mounted on the block and have leads that leave to other wiring in the truck. This lets the injector modules "push" against the wiring in the truck, making the poorly grounded engine block pump up and down with pulse voltages. Since the exhaust system bolts to the engine with a direct conductive path through the turbocharger, the tailpipe is actually excited just like an antenna.

I cured ALL of the remaining noise with the addition of one more ground strap between the exhaust and the frame just ahead of the muffler.

You can see the widening of the pipe as it expands into the muffler, with the fuel tank and transfer case in the background. The strap grounds under a nut used for the exhaust hanger. As before, I tin the leads and use stainless star washers to help maintain connections. Notice I clamped the strap under the factory exhaust joint clamp.

I now have absolutely no noise at all from the diesel engine on any band.

since July 5 2004

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My Shack

[Home]

FT1000D

IC751A's

Modified R4C's (phase locked to each other)

AL800 on 6 meters

AL1200 for FT1000

AL1500 for IC751A

Skin depth and Conductors

[Home]

The following short article is about Litz wire, braid, and skin depth. It is still being expanded.

The best explanation of skin depth I have seen is in Circuits and Networks (by Koehler) on page 196:

conductor with return path at a distance equal to several times the diameter of the conductor. Assume that an emf is applied between the two ends of the conductor and further assume that the conductor is divided into many concentric cylinders of equal cross-sectional area. Now if the current density were uniform over the entire cross section, the total current in each cylindrical shell would be the same. The magnetic flux encircling the innermost cylinder would be larger than that of the outermost cylinder. Since the voltage drop along each of the cylinders is made up of a resistance drop (IR) plus a reactance drop (jIX), the reactance drop along the innermost cylinder would be higher than that of the outermost cylinder because of the greater flux. Obviously this cannot be true because the voltage drop across all cylinders must be the same, i.e., the impressed voltage. Consequently, the current density changes in such a manner as to equalize all voltage drops. This results in a higher current density toward the outside surface of the conductor. This effect, called skin effect, is a function of the frequency, permeability, conductivity, and diameter of the conductor. It causes the conductor to have a higher ac resistance than dc resistance because it effectively reduces the size of the conductor by virtue of the decrease in current density from the outside to the inside portion of the conductor. The current not only decreases in magnitude but also changes in phase with reference to the current at the surface. At high frequencies, dependent upon the size of the conductor and its electrical properties, most of the current may flow only in a very thin outer cylinder.

An exact mathematical analysis of current distribution in a cylindrical conductor shows that the current density is given in terms of Bessel functions of a complex argument that involves the distance from the center divided by δ where $\delta = 1/\sqrt{\pi f \mu \sigma}$; μ is the permeability and σ is the conductivity. When δ is very small compared to the radius of the conductor the ac resistance of the conductor is substantially the same

the conductivity. When δ is very small compared to the radius of the conductor the ac resistance of the conductor is substantially the same as the dc resistance of a conductor with a cross-sectional area equal to $2\pi r\delta$, where r is the radius of the conductor.

When δ is not small compared to r, it cannot be assumed that the entire current flows in a cylinder of radius r and thickness δ. For large values of δ/r, an exact solution* for the current distribution results in the curve of Fig. 8-3, where R_{ac}/R_{dc} is shown as a function of r/δ. This curve indicates that when $r/\delta \leq 1$, the ac and dc resistances are substantially equal. Thus if it desired to have the resistance of a conductor

So we see, skin effect occurs because the outer areas of a conductor have less magnetic flux surrounding them than the inner layers. While we often consider the effect on materials without weaves, think about how this affects woven conductors like braiding or Litz wire. Any individual conductor has HIGHER impedance when it moves into the center areas, and if current finds an alternate path it moves out to the lower-impedance outer layer.

If strands touch with high resistance connections, it effectively adds unnecessary resistance to the path. The fact the surface is rough also DECREASES effective surface area for current for a given occupied physical area, because there are non-conductive air gaps in the occupied area!

The increased resistance (and impedance) is why braiding should never be used for high frequency high current applications unless it is oversized to compensate for increased impedance and dissipation. Parallel stranding decreases effective current carrying surface area at high frequencies, but at least avoids the weave problems.

As a matter of fact the weave is what causes most of the loss increase when typical coaxial cables are contaminated by water. The problem occurs, even after the cable is dry, because the hundreds of weave contact points corrode and make poor connections, increasing the resistance. Something similar happens when we "unpack" a braid from a jacket. The pressure between weaves decreases, and series resistance (and impedance) increases!

We could insulate strands, and once we establish a basic loss in conductivity caused by replacing useful conductor area with insulation we would find the impedance slope much flatter with frequency. If the strands are insulated (Litz wire) and frequency is increased, eventually we reach a point where the conductor resistance increases and we are much worse off than we would be with braided or stranded wire. The upper limit point is generally around 1 MHz for Litz wire.

From Radio Engineering by Terman:

Litz Wire.—The effective alternating-current resistance of a conductor can be made to approach the direct-current resistance at low and moderate radio frequencies, by forming the conductor from a number of strands of small enameled wires connected in parallel at their ends, but insulated throughout the rest of their length, and thoroughly interwoven. If the stranding is properly done, each wire will, on the average, link with the same number of flux lines as every other wire, and the current will divide evenly among the strands. If at the same time each strand is of small diameter, it will have relatively little skin effect over its cross section, so all of the material is equally effective in carrying the current. Such a stranded cable is called a *litz* conductor.

Practical litz conductors are very effective at frequencies below about 1000 kc, but as the frequency becomes higher the benefits disappear. This is because irregularities of stranding, and capacitance between the strands, cause a failure to realize the ideal condition at very high frequencies.

In reality, the real advantage of Litz wire is NOT reduced resistance per unit length for a given diameter. The advantage is less slope in resistance with frequency and reduced eddy current when the conductor is in a multi-layer coil or transformer. The individual strands are like laminations in a transformer core, and below a certain frequency they greatly reduce eddy currents by decreasing the "short circuit" path distance for magnetic flux induced currents that are not in the normal current flow direction.

My own measurements have shown noticeably lower ESR (equivalent series resistance) for a given conductor size when using a ***SOLID*** wire as compared to any Litz wire sample I have tried above 300kHz (the lower limit of my main vector network analyzers).

For a given wire diameter and form factor, I've always been able to achieve higher Q with solid round or ribbon conductors.

Also see mobile and loaded antennas.

stacking broadside collinear

[Home]

Popular rumor is gain doubles each time elements are doubled. In other words, adding a second vertical above another vertical would add 3dB gain. Adding two more after that (total of four) produces 6dB gain over a single element.

The same concept is often applied to stacked Yagi or dipole antennas.

In neither case is it true. Gain comes from forcing nulls in areas of high radiation. This removes energy from the null areas, and that energy moves to enhance gain in other directions. This effect is sometimes called ***pattern multiplication***.

Before antenna modeling software was commonly available, we often used pattern multiplication to estimate patterns and gain. It still is a useful tool.

The graphs below, from Jasik's Antenna Engineering Handbook, shows the gain of various couplets or elements when placed either end-to-end (Collinear) or parallel above each other (stacked broadside).

Collinear Gain

First we have the end-to-end element placement gain.

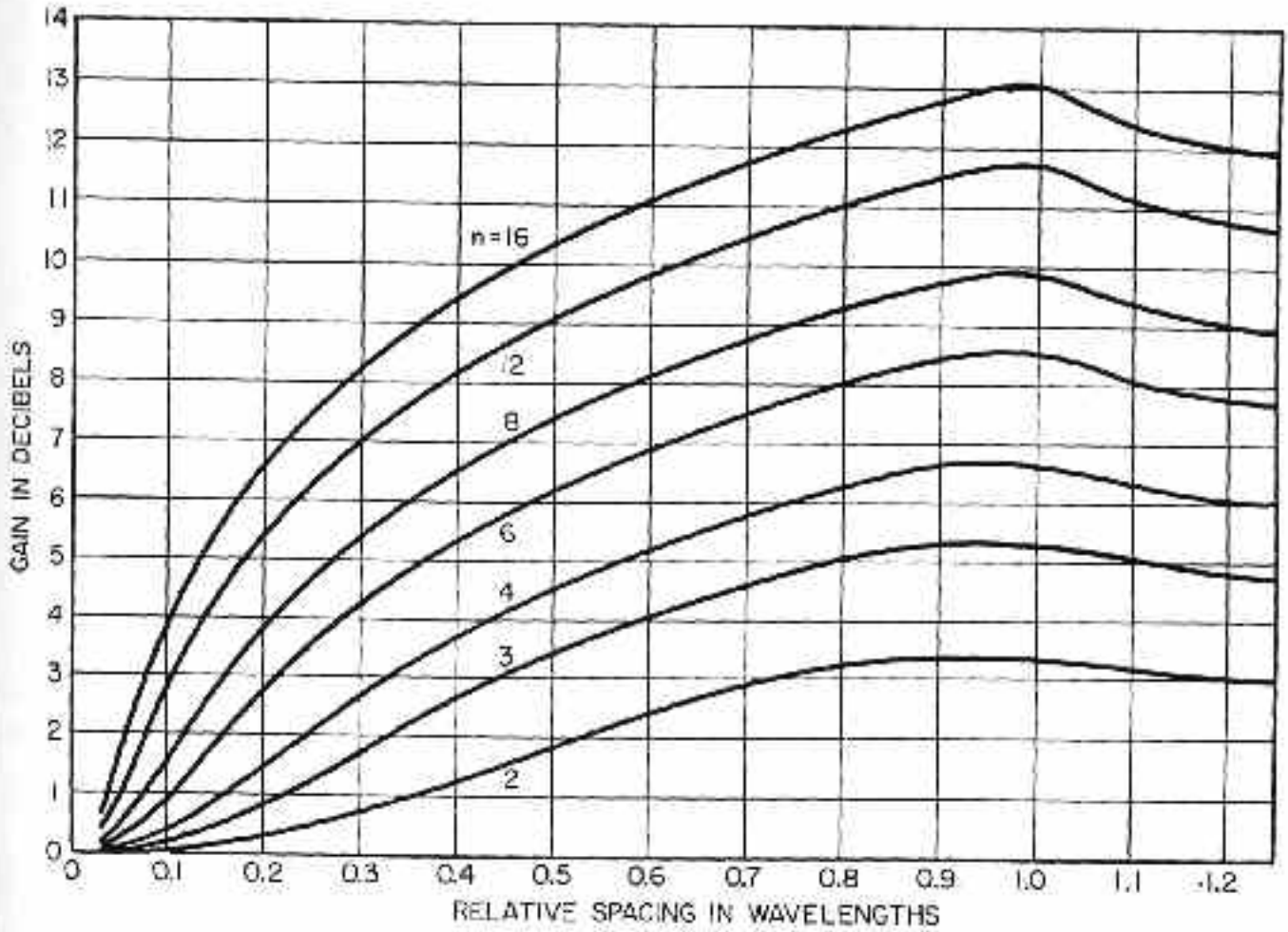


FIG. 5-22. Gains to be expected from a collinear (omnidirectional) array of short-dipole elements. The gain is relative to that of a single element.

We can see the gain for two elements peaks at .9 wavelength spacing. This spacing is the current maximum spacing of the elements, NOT the end spacing. With a 1/2 wl dipole in each element, the end-to-end spacing would be .9 - .25 -.25 = .4 wavelengths. The overall array length would be .9 + .25 + .25 = 1.4wl

With two dipoles end-to-end the center-to-center spacing of current maximums would be .25+.25=.5wl. The absolute maximum gain would be found on the graph above at the crossing of the vertical .5 relative spacing line and intersection of curve 2 (two elements). The gain would be less than 1.9dB in ANY collinear antenna.

We would achieve 3dB at about .71-.25-.25= .21wl tip-to-tip spacing, or 1.21wl total collinear element array length.

To double gain again (adding 3dB more) the array would have to be four elements with .75wl center-to-center spacing in elements! The array would be 3*.75=2.25wl (this is number of elements minus 1 multiplied by required spacing) for the total spacing of four elements). The end-elements would extend .25 wl each from the current maximum, so the array length from end-to-end would be

2.25+.25+.25=2.75wl.

To go from three to six dB requires we change antenna length from 1.21wl to 2.75wl!

We cannot simply double length to double gain! That concept is wrong.

Note we can get more than 6dB gain by using more than .75wl element center-to- element center spacing. We really should have .95wl element center-to element center spacing, making the array N-1*S + El = L where N=number of elements, S=spacing, El = element length. The result is an array of four elements would have optimum gain of 6.7dB with a length of 3.35wl.

The longer the array is, the wider the individual elements should be spaced for optimum gain.

Broadside

Broadside gain applies to elements that are parallel and one above the other. This could be Yagis or dipoles. This is the ***OPTIMUM or maximum gain***, not the actually gain you might have.

Optimum broadside stacking distance increases with more directive elements, so a pair of three-element Yagis stacked requires wider spacing than a pair of dipoles, and less maximum stacking gain is possible.

Here is the optimum gain graph for dipoles:

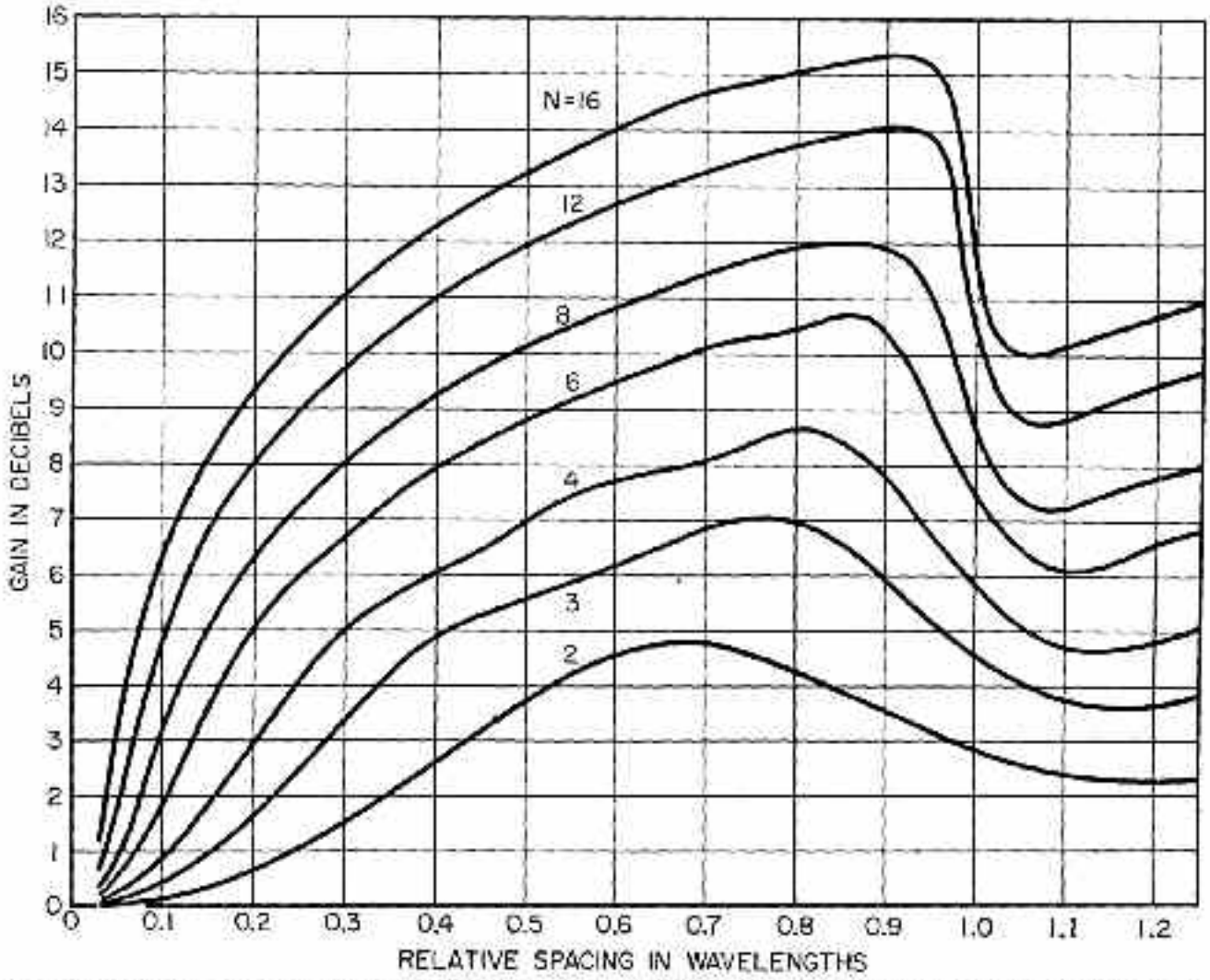


FIG. 5-10. Gains* to be expected from a single array (no reflector) of broadside elements.

You can see maximum gain occurs at .675wl stacking height. The stacking gain is 4.8db, not 3dB as we see claimed claim. Again the more elements, the wider the stack spacing must be between elements.

I hope these graphs help dispel the myth that doubling element numbers doubles gain! It just doesn't work that way.

A Practical Application?

Combination Broadside/Collinear/ End-fire

If all goes as planned, a new 3.5-10MHz array will be installed later this year in the SE USA using four 80-meter 1/2wl elements fed with open wire line with optimum spacing in broadside-endfire. This array will be 300ft tall and 400 feet wide. A reflector grid of wires will be spaced 1/8th wl behind the array.

The 80 meter gain is 4.8dB broadside + 3.2dB collinear for a total of 8dBd. A screen reflector adds 3dB gain, for a total of 11dBd gain on 80 meters. This is 11dB over a dipole at the same mean height! With ground reflection gain the total dBi gain is about 18dBi.

On 40 meters, the same array will produce 21dBi (or 14dB gain over a dipole at the same mean height). In contrast, a typical 3-element full size Yagi on 80 meters would have about 7dBd gain. The curtain is roughly like four stacked three-element Yagis on 80, and eight stacked on 40.

Curtain pattern can be slewed in both E and H planes to focus on the target. Such arrays are actually practical, and have been used for years in SW BC work.

W4AN

[Home]

Trust Fund Information

There have been many information requests for the trust fund for Graeme and Erik, Bill and Dana's sons. The family is overwhelmed by the outpouring of kind words by fellow amateurs. If you would like to contribute to the W4AN Memorial Fund for Bill and Dana's children, the trust fund account information is:

Account number 20005913

W4AN Memorial Account for Graeme and Erik Fisher

North Atlanta National Bank

10500 Old Alabama Road Connector

Alpharetta, GA 30022

Bank Phone (678-277-8400)

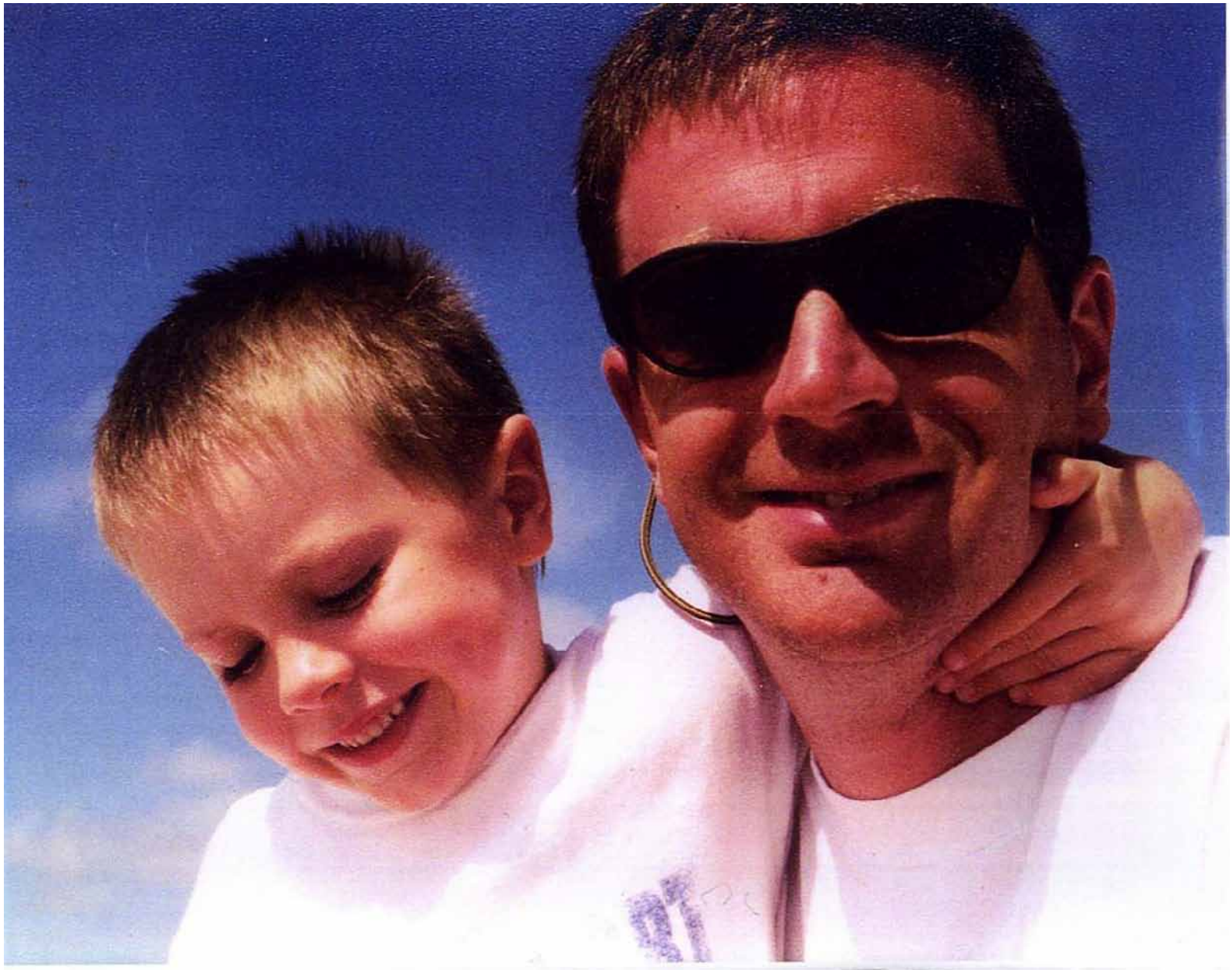
If you wish to make a contribution, any amount will help. Make checks payable to:

W4AN Memorial Account.

You can include the account number "**20005913**" or "**Graeme and Erik Fisher**" in the "for" line.

I will post more information after the Dayton Hamvention rush is over and I recover. Briefly, the basic plan is that Graeme and Erik will be able to draw from this fund for worthwhile activities like school, a first home, or some valid emergency. Having a headache, buying new mag wheels or a dirt bike would not be a good reason. College tuition, books, or housing would be. It will be a 50/50 split unless one doesn't require funds. If neither requires funds, then it will go to some other similar circumstance involving an amateurs young children.

The fund is administered by several well-known members of the amateur community. All of this will be released in final form with more detail a short while after Dayton. Sorry, but anyone who works in the industry knows what a huge time-sink Dayton is!



Graeme and Bill Fisher

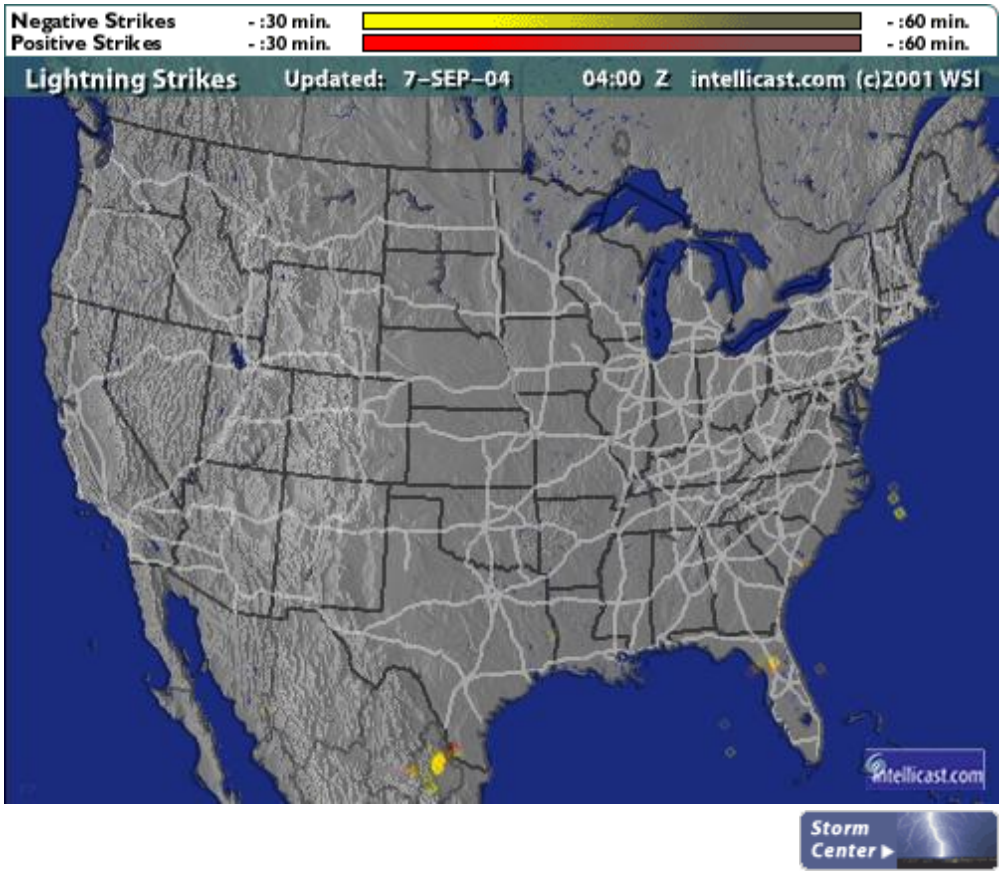
Active Tropical Systems! Click here for details ▶

Active Watch/Warnings

Site Spotlight

National Lightning

National Lightning



Storm Center ▶

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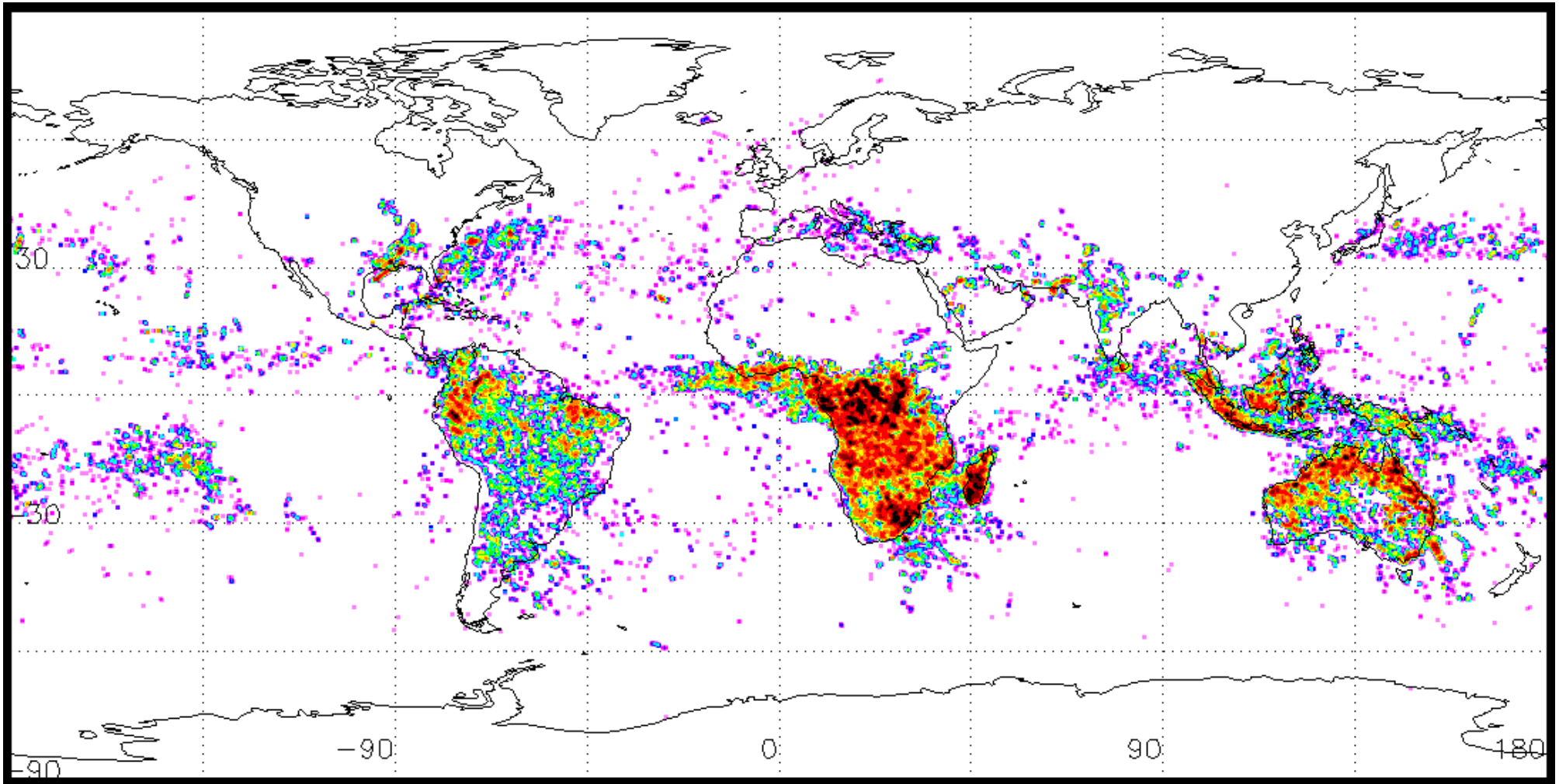

Orbits	799
Areas	27117
Flashes	137810
Groups	683597
Events	1463457

(Created : 02/14/100)

1 2 3 4 5 >5 >10 >15 >25 >50 >100 >150

Flash scale

1998 December, 1999 January, 1999 February

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Look in on part of SEI's Laboratory

More information on shortwave receivers & stations from Passport to World Band Radio

Receiver Test Data

Device Under Test	Noise Floor	AGC Thrshld	dB	100kHz Blocking	Sensitivity	LO Noise Spacing	kHz	Front End Selectivity	Filter Ultimate	Dynamic Range Wide Spaced	kHz	Dynamic Range Narrow Spaced	kHz
Drake R-4C/CF-600/6	-138 dBm	0.7 uV	3	130dB	0.15 uV	135 dBc	10	A-Preselector	130dB	85dB	20	84 dB	2
AOR AR-7030	-122/-128*	2.2	3	130	0.5/0.22	130	10	D Hi /Lo Pass	90	100	20	82	2
Atlas 350-XL	-131	1.0	11	117	0.2	125	4	C Band Pass	95	81	20	81	2
Kenwood TS-830/YK88	-129	1.5	3	122	0.1	114	2	C Preselector	85^	84	20	81	2
Added 2/27/04 Ten-Tec Omni VI+	-135	0.7	3	145	0.2	137	20	B Band Pass	80	97	20	80	2
Yaesu 901-DM	-135	1.6	3	124	0.15	109	2	C Preselector	85	87	20	80^	3
Collins R-390A	-137	N.A.		130	0.2	130	2	A+ Trk Presel	85	81	20	79	2
Ten-Tec Corsair	-131#	0.1	14	130	0.2	132	5	C Band Pass	90	93	20	79	3

Icom IC-720A	-137	1.6	3	138	0.15	117	10	C 0.5 Octave	80	93	50	78	3
Kenwood TS-820S	-137	0.4	3	115	0.2	125	10	C Preselector	80	79	20	78	3
JRC NRD-515	-138	3.5	4	103	0.1	118	10	C 0.8 Octave	80	95	20	77^	2
Ten-Tec Omni V	-134	1.2	6	135	0.18	134	10	C Band Pass	100	89	20	76	2
Atlas 210/215X	-120#	N.A.		123	0.5	N.A.		C Band Pass	95	76	20	76	2
Drake R-7	-135/-140*	1.0/0.4*	3	145	0.28/0.15*	114	10	B 0.5 Octave	85	97	100	75	2
Drake TR-7	-134	1.3	3	146	0.5	116	10	B 0.5 Octave	90	99	100	75	2
Heath SB-104	-123	N.A.		92	0.5	N.A.		C Band Pass	75	79	20	75	4
WJ HF-1000	-129/-136#*	0.11	3	123	0.23/0.13*	115	10	D Wideband	80	99	20	75	5
Ten-Tec Omni-B	-136	0.2	25	129	0.15	130	10	C Preselector	80	87	20	74	2
Icom IC-730	-140	1.5	3	135	0.1	118	10	C 0.5 Octave	80	92	50	74	3
Kenwood R-820S	-125	4.0	3	125	0.35	123	10	C Preselector	75	74	20	74	4
Collins 75-S3B	-146	1.1	15	122	0.1	120	4	B+ Preselector	85	88	20	74	2
Icom IC-781	-127/-138*	2.0/0.5*	3		0.5/0.18*	129	10	B 0.5 Octave	90^	94	20	73	2
Stock 781	-128/-135	2.4/0.7	3	131	0.5/0.22			B 0.5 Octave	90	98	20	78	2
781 Pin Diodes	-126/-134	3.6/1.15	3	134	0.54/0.21			B 0.5 Octave	90	98	20	72	2
Kenwood TS-930S	-135	2.0	3	143	0.15	115	10	B- 0.5 Octave	80^	86	20	73	3
Icom IC-701	-129	5.5	6	130	0.3	125	10	C Band Pass	75	81	50	73	4
Collins 75S-3C	-141	1.3	12	121	0.14	120	4	B+ Preselector	95	85@	20	72	2
JRC NRD-525	-132#	0.9	3	123	0.2	120	10	B Trk Presel	65	95	50	72	5

Drake R-8	-128/-131#*	0.6/0.3	3	130	0.25/0.18*	115	10	C 0.5 Octave	75^	90/85*	20	71	5
Icom IC-R72	-127/-135#	3.1/1.2*	3	129	0.28/0.11*	122	10	C 0.5 Octave	75	87*	20	71	5
Icom R-9000	-131#	0.8	3	129	0.15	128	10	B 0.5 Octave	90	93	20	71	5
Added 2/27/04 Elecraft K2	-135	2.6	15	118	0.22	123	10	C Band Pass	80^	95	20	70	2
JRC NRD-535	-135#	0.9	3	114	0.1	117	10	B Trk Presel	70	92	50	70	5
Kenwood TS-830S	-136#	0.9	3	122	0.1	113	2	C Preselector	80	84	20	70	3
Icom IC-761	-131/-139*	2.0/0.7*	3	145	0.4/0.17*	129	10	B- 0.5 Octave	90^	87	20	70	2
Lowe HF-150	-126#	0.7	3	126	0.3	113	10	F No Bandpass	75	84	20	69	5
Kenwood TS-430S	-136#	0.6	3	134	0.1	102	10	C 0.5 Octave	70	78	20	69	5
JRC NRD-545	-130#	2.0	6	127	0.2	118	10	B 0.5 Octave	65	96	100	66	5
Signal/One CX-11A	-122#	0.6	17	109	0.6	119	50	C 0.5 Octave	105	90	50	68	5^
Kenwood TS-180S	-139	0.9	3	115	0.15	120	10	C Preselector	80	70	20	68	3
Drake TR-4C	-124#	1.2	3	105	0.4	130	10	C Preselector	80	74	20	68	2
Icom IC-735	-126/-133*	1.5	12	135	0.35/0.18*	123	10	C 0.5 Octave	90^	83	20	68	2
Added 1/21/04 Icom IC-R75	-123#/-130*	3.5/1.3* uV	3	119	0.5/0.2*	109	10	B 0.5 Octave	80	95	50	67	5
Drake SW8	-127#	0.9	3	125	0.32	113	10	N.A.	70	92	20	67	5
Racal 6790 GM	-128	0.3	1	145	0.7	130	10	D Broadband	85	95	20	66	2
Lowe HF-235	-126#	0.8	3	129	0.35	117	10	D Octave	80	71	20	66	5

AOR AR3030	-131#	2.0	10	130	0.16	117	10	C 0.5 Octave	85	90	20	66	5
Yaesu FRG-100	-133#	0.9	3	127	0.13	112	10	C 0.5 Octave	70	99	50	65	5
Kenwood R-5000	-131#	0.4	3	134	0.2	120	10	C 0.5 Octave	80^	86	20	65	5
Added 2/27/04 Palstar R-30	-123	2.6	3	130	0.35	116	10	C Octave	90	88	20	64	5
Yaesu FRG-7700	-130#	1.3	3	123	0.2	100	10	D Octave	65	83	50	64^	5 t
Kenwood R-1000	-130#	0.9	3	119	0.2	107	10	D Octave	70	76	20	64^	3 t
Heath SB-303	-134	N.A.		104	0.5	N.A.		C Preselector	70	66	20	64	4
Collins KWM-380	-127#	1.1	5	123	0.3	99	10	B 0.5 Octave	70^	94	50	64^	2
Icom IC-751	-127/-133*	6.3	3	138	0.4/0.2*	127	10	B- 0.5 Octave	90^	84	20	64	2
Kenwood TS-520	-139	N.A.		116	0.15	N.A.		C Preselector	70	63	20	63	3
Yaesu FT-One	-135	1.0	3	130	0.2	99	10	C 0.5 Octave	80^	91	50	63^	2
Collins 75-S3 Wing	-145	1.0	14	105	0.1	N.A.		B Preselector	75	75	20	63	3
JRC NRD-93	-141	1.6	3	128	0.15	133	10	A+ Trk Presel	80	94	20	63	2
Yaesu FT-980	-136	1.8	3	140	0.12	106	10	C 0.5 Octave	62^	96	50	63	2
Icom IC-R70/R-71A	-129/-135*	3.1/1.4*	3	132	0.4/0.2*	128	10	B- 0.5 Octave	90^	86	20	62	3
Grundig Satellite 700	-127#	1.6	3	106	0.3	118	10	N.A.	85	76	20	62	5
KWZ-30	-130	1.0	3	120	0.8	118	10	D	80	100	20	60	5
Collins 51S1	-134	1.0	7	117	0.13	146	10	A Trk Presel	100	84 <>	100	60	5

Icom R-8500	-135#	0.45	3	132	0.11	131	10	B 0.5 Octave	75	81	20	59	5
Yaesu FT-101E	-141	N.A.		102	0.15	N.A.		C Preselector	70	60	20	59	3
Drake R-4C Stock	-139	0.7	3	130	0.15	135	10	A-Preselector	70	85	20	58	2
Yaesu FT-757	-120/-134*	1.6	3	130	0.7/0.15*	109	10	C 0.5 Octave	70^	86	20	56	3
Ten-Tec 340	-123/-133#*	0.5/0.13	3	109	0.4/0.14	113	10	B 0.5 Octave	70	93	100	46	5
Kenwood R-2000	-130#	1.4	3	115	0.15	105	10	D Octave	70	71	20	45	5
Kenwood R-600	-130#	0.8	3	109	0.2	99	10	D Octave	65	68	20	F.L.	5
Yaesu FRG-8800	-132#	0.6	3	122	0.18	N.A.		D Octave	70	87	20	F.L.	5
AOR 5000	-124/-130#*	0.9/1.8	3	118	0.2/0.35	103	10	B 0.5 Octave	60	58	50	^	5

Notes on Receiver Table Legend:

One (1) dB blocking (gain compression) test done at 100 kHz to eliminate phase noise interaction.
N.A. = Data not available.
# = Measured with SSB filter.
***** = Built-in preamp actuated.
F.L = Filter limited, no measurement was possible due to signal leakage around filter.
t = Readings would have been lower if 2-kHz spacing had been possible.
^ = Measurement was phase-noise limited rather than intermodulation limited.
@ = Dynamic range is 90 dB at 100 kHz spacing.
<> = 20kHz dynamic range is 66dB.

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Up: W9CF Links

Spectral Analysis of a CW keying pulse

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October 6, 2001

1 Introduction

This problem was brought to my attention by Tom Rauch, W8JI, who, in his note to me, had described his experience and correctly pointed out all the main features that govern the bandwith. These notes are an expanded version of my reply giving the mathematical explanation.

To fit the most CW signals into the available spectrum, we need to limit the bandwidth taken up by the signals. It is therefore useful to see how the energy in a dot or dash pulse is distributed around the carrier frequency. Here I give some notes on how to make this analysis. The main result is that the spectrum for many keying shapes is given by the product of the spectrum of a square pulse times the spectrum of the slope of the rise and fall behavior of the pulse.

It seems from my experience reading morse, that the rise time should be the main factor in producing code that can be read by ear comfortably. Since the rise time dominates the bandwidth for the usual CW signal, the analysis shows that to get a nearly optimal bandwidth to rise time, the keying pulse shape should have a gaussian slope.

In the next section I review basic Fourier analysis of amplitude modulation. I then calculate the spectrum of a pulse with an exponentially shaped rise and fall as would be produced by simple RC networks. The results suggest the more general analysis in the following section, with the conclusion that a pulse with gaussian slope, i.e. error function rise and fall shapes, will have an optimal bandwidth and rise time.

It seems likely that all of this would have been worked out by radio engineers in the early 1900s when CW signals were first employed.

2 Fourier analysis for amplitude modulation

To analyze the spectrum generated by keying a transmitter let's look at a single ``dot.'' If we imagine we have a

carrier with angular frequency ω_0, and we amplitude modulate it with an envelop A(t), we get the amplitude of the signal from the transmitter is

$$f(t) = A(t)\cos(\omega_0 t) \tag{1}$$

To calculate the energy spectrum we Fourier transform this to get

$$\tilde{f}(\omega) = \int_{-\infty}^{\infty} dt A(t)\cos(\omega_0 t)e^{i\omega t} = \frac{1}{2}\left[\tilde{A}(\omega-\omega_0)+\tilde{A}(\omega+\omega_0)\right] \tag{2}$$

where

$$\tilde{A}(\omega) = \int_{-\infty}^{\infty} dt A(t)e^{i\omega t}. \tag{3}$$

In the usual case, the modulation $A(t)$ contains frequency components much smaller than the carrier frequency. Therefore the $\tilde{A}(\omega+\omega_0)$ is negligible and can be ignored. The energy spectrum of the amplitude modulated signal is therefore given by

$$P_0(\omega) = \frac{1}{4\pi}|\tilde{A}(\omega-\omega_0)|^2 \tag{4}$$

and since $A(t)$ is real, $\tilde{A}(\omega) = \tilde{A}^*(-\omega)$, and the sidebands are symmetric around the carrier frequency.

It is convenient to write

$$P(\omega) = P_0(\omega+\omega_0) \tag{5}$$

so that ω here is the frequency difference from the carrier, so that $P(\omega)$ gives the energy distribution for an

angular frequency of ω from the carrier angular frequency. I will call $P(\omega)$ the sideband energy density.

3 Application to Keying Bandwidth

To get an explicit result, I'll assume an explicit form for a keying waveform. A simple form where the Fourier transforms can be calculated analytically is the case where the wave builds up exponentially (as in the usual RC circuit) to the carrier value when the key is pressed, and then decays exponetially to zero when the key is released. That is

$$A(t) = \begin{cases} 0 & t < 0 \\ 1 - e^{-t/\tau} & 0 < t < T \\ (1 - e^{-T/\tau})e^{-(t-T)/\tau} & t > T \end{cases} \tag{6}$$

where T is the keying pulse width and τ is its time constant. A plot of this waveform is shown in figure 1 for the cases where τ is 4 milliseconds, and T is 20 and 50 milliseconds.

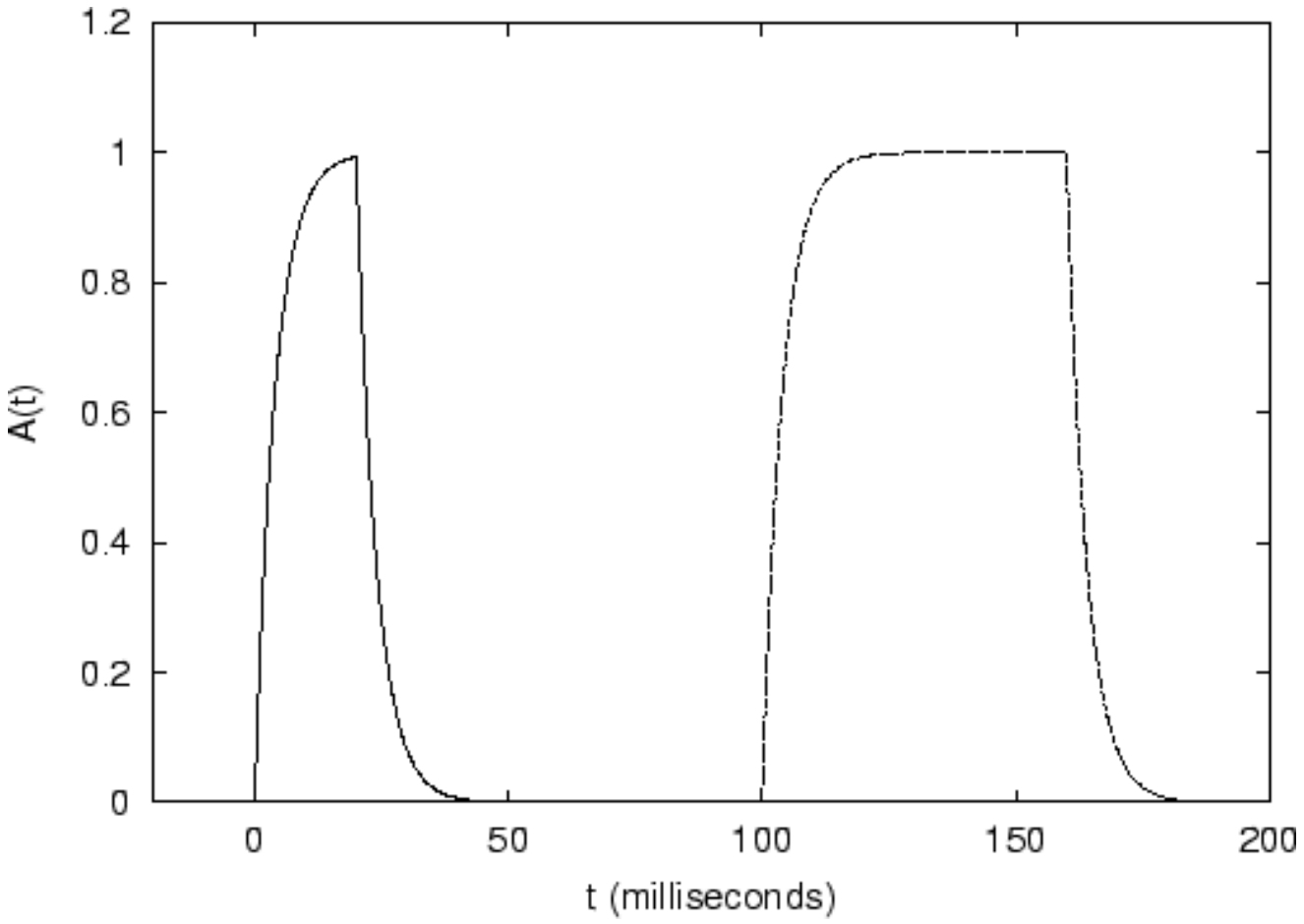

Figure 1: The exponential keying waveforms for a time constant τ of 4 milliseconds and durations of 20 and 50 milliseconds. The 50 millisecond pulse begins at $t = 100$ milliseconds to separate it from the 20 millisecond pulse.

The fourier transform integral of the amplitude is straightforward and gives

$$\tilde{A}(\omega) = -\left[e^{i\omega T} - 1\right] \frac{1}{i\omega(i\omega\tau - 1)} \tag{7}$$

so that the sideband energy density) becomes

$$P(\omega) = \frac{1}{\pi} \frac{\sin^2\left(\frac{\omega T}{2}\right)}{\omega^2} \frac{1}{(1 + \omega^2\tau^2)} \,. \tag{8}$$

So the sideband energy density has two factors. If we measure it in dB relative to some fixed value, we add the logarithms of the factors. The only dependence on T comes from the first factor. The $\sin^2$ function is always less than or equal to 1, so this will subtract from the other factor which only depends on τ the time constant which determines the rise time.

The main features of the sidebands will therefore be given by the rise time, while the length of the pulse will modify those features somewhat.

Figures 2 and 3 show the energy density in dB referenced to the carrier energy density for T of 20 milliseconds. The curves are plotted together in figure 4. Each curve is plotted versus frequency $f = \omega/2\pi$.

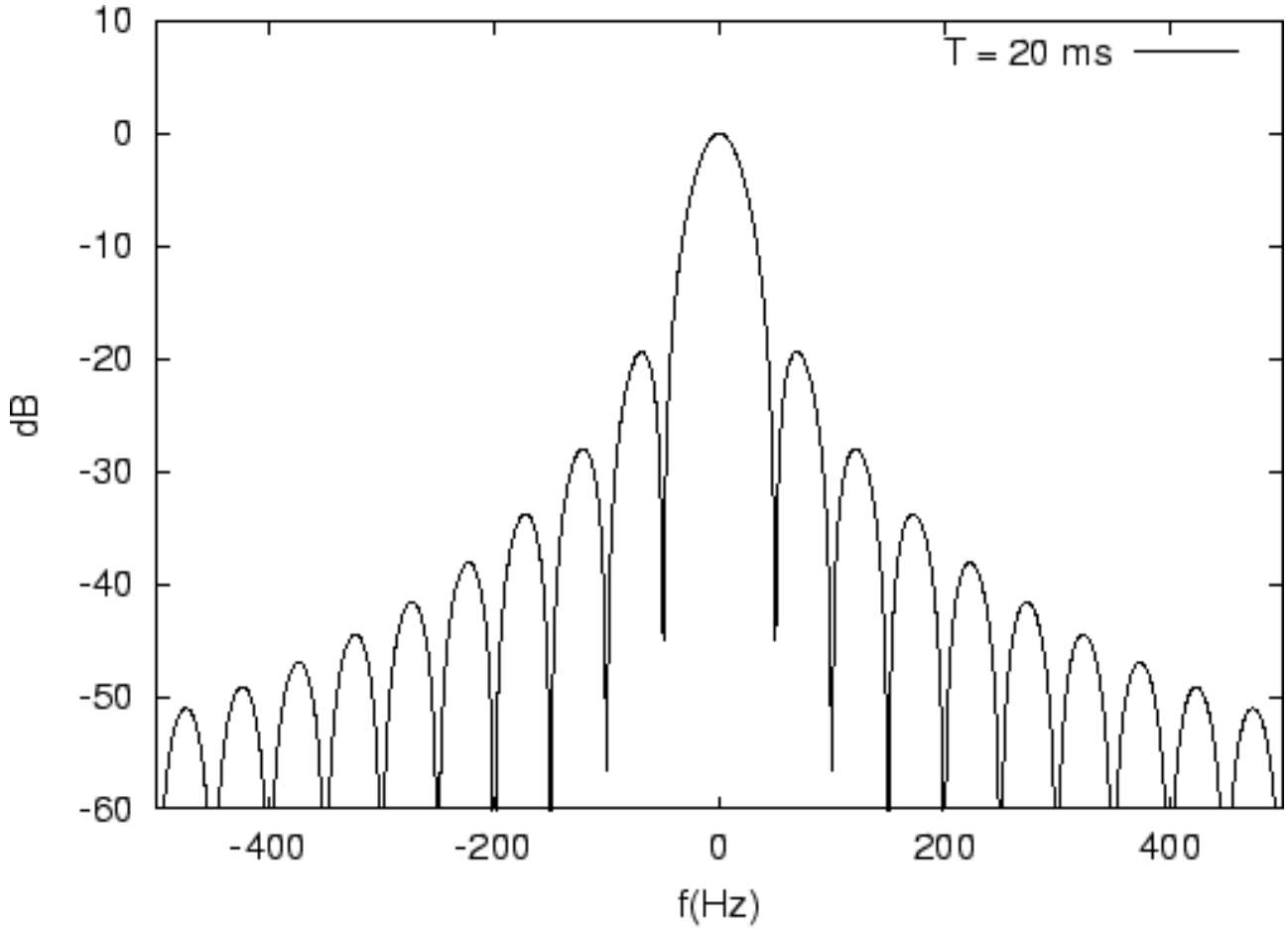

Figure 2: The sideband energy density for an exponential keying wave form with $\tau = 4$ milliseconds and T of 20 milliseconds. The carrier energy density is set to 0 dB.

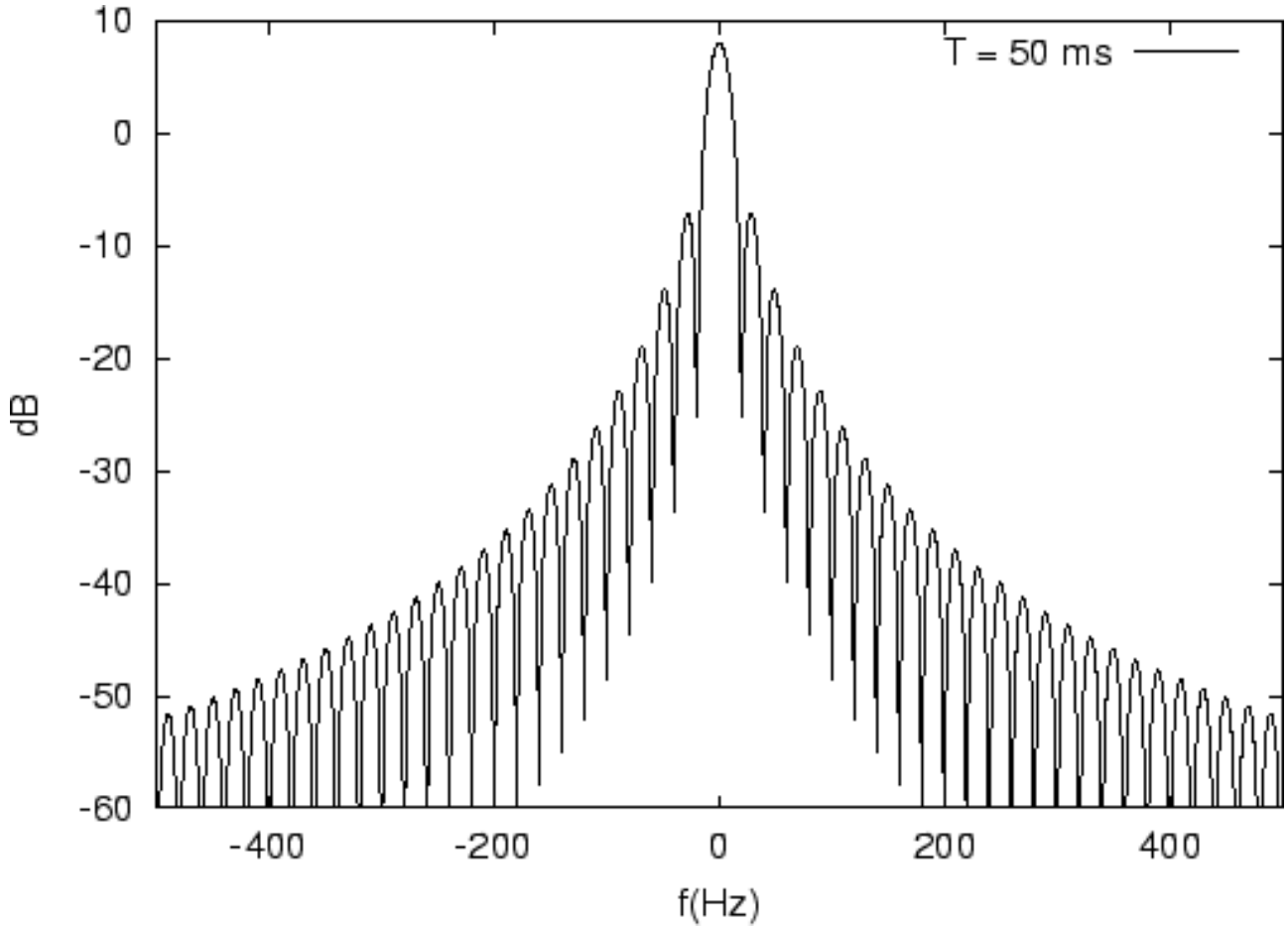

Figure 3: The sideband energy density for an exponential keying wave form with $\tau = 4$ milliseconds and T of 50 milliseconds. The scale is the same as figure 2.

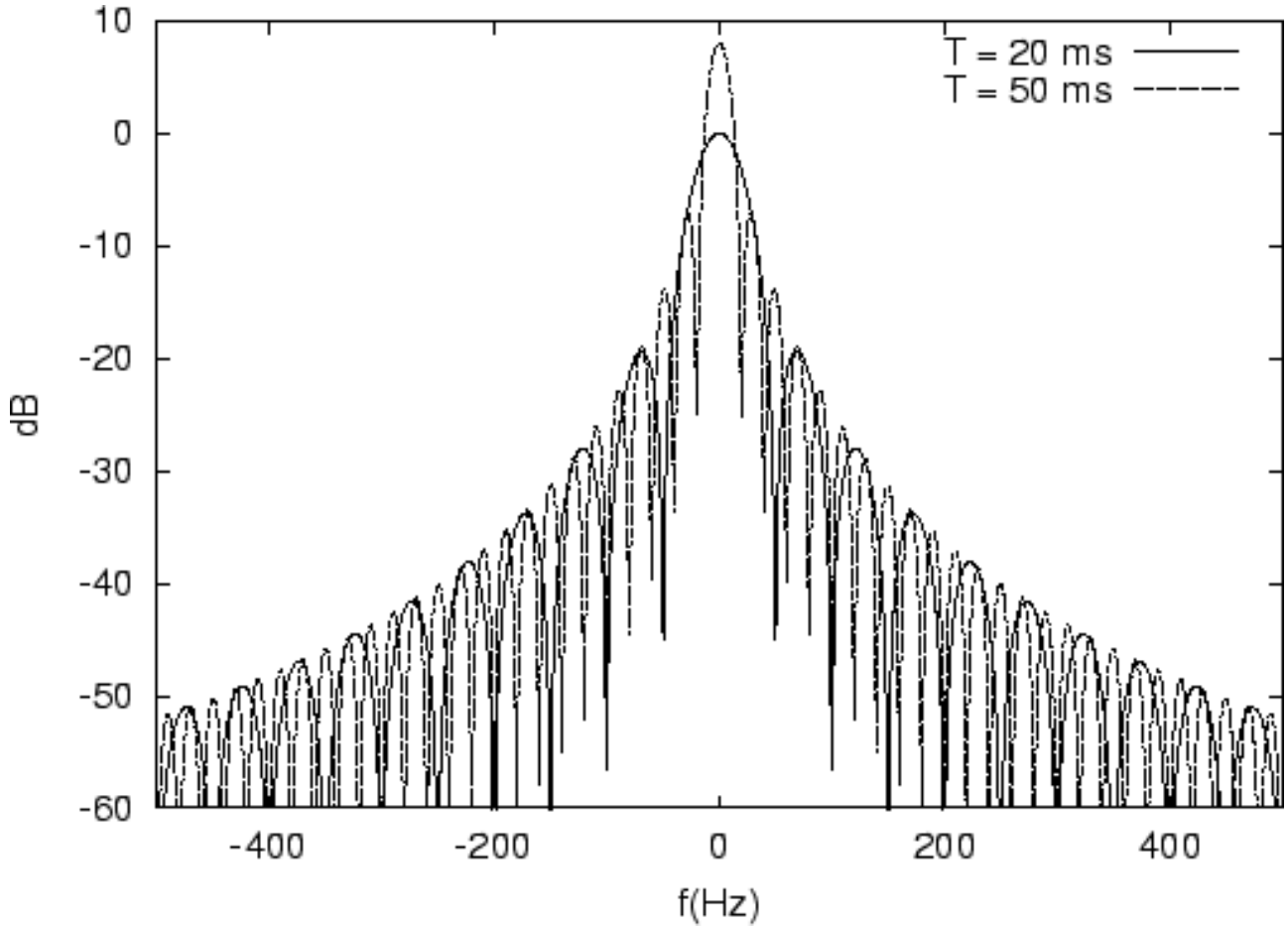

Figure 4: The plots of figures 2and 3combined.

Notice that the differences between the 20 millisecond and 50 millisecond pulses are first the energy near the carrier frequency is larger for the longer pulse as needed since it has about 2.5 times as much energy, and second the ``ringing" has more oscillations for the longer pulse as expected. The sidebands fall off 12 dB per octave once we are at frequencies beyond about $1/\tau$.

The effect of the keying speed on the bandwidth as long as the rise time is small compared to pulse length is the change in shape of the central peak. It does get narrower for slower keying and wider for faster keying, however, the keying speed does not effect the overall bandwidth.

4 General Pulse Shape

I can write a general pulse shape as

$$A(t) = E_r(t+T/2) - E_f(t-T/2) \tag{9}$$

where E_r and E_f describe the rising and falling edges of a pulse and are positive functions that go to zero at

large negative t and to 1 at large positive t. The Fourier transform of A is the sum of the transforms of the two terms. If the rising and falling edges have the same form, we can write

$$A(t) = E(t+T/2)-E(t-T/2) = \int_{-\infty}^{\infty} dt'[\delta(t+T/2-t')-\delta(t-T/2-t')]E(t') \quad (10)$$

and integrating by parts gives the result

$$A(t) = -\int_{-\infty}^{\infty} dt' S(t-t')\frac{dE(t')}{dt'} \equiv -\int_{-\infty}^{\infty} dt' S(t-t')E'(t') \quad (11)$$

where $S(t)$ is a square pulse of width T,

$$S(t) = \begin{cases} 1 & |t| < \frac{T}{2} \\ 0 & |t| > \frac{T}{2} \end{cases} \quad (12)$$

and I have defined E' to be the derivative of E.

Since $A(t)$ is written as a convolution, its Fourier transform is now the product

$$\tilde{A}(\omega) = \tilde{S}(\omega)\tilde{E}'(\omega)\,. \quad (13)$$

For keying wave forms this has the nice interpretation that the spectrum is given by the spectrum of a square pulse of length T multiplied by a factor that is the Fourier transform of the slope of the rise and fall wave form.

Except for an unimportant change of the zero of time, the exponential case calculated above can be written as

$$E(t) = \begin{cases} 0 & t < 0 \\ 1 - e^{-t/\tau} & t > 0 \end{cases} \quad (14)$$

The Fourier transforms of these are

$$\begin{aligned} \tilde{S}(\omega) &= \frac{2\sin\left(\frac{\omega T}{2}\right)}{\omega} \\ \tilde{E}'(\omega) &= -\frac{i}{1+i\omega\tau} \end{aligned} \tag{15}$$

and the sideband energy density is exactly as before.

As we saw for the exponential case, the bandwidth is dominated by the rise and fall time. Therefore it seems reasonable to try to optimize the rise and fall waveform of the keying pulse. In terms of the function $E'(t)$ whose integral is the rising and falling wave form, we want to simultaneously make its width in real time and in frequency small. One measure of this is the product of $\Delta\omega$ and Δt where they are defined as the variances in frequency and time

$$\begin{aligned} (\Delta t)^2 &= \int_{-\infty}^{\infty} dt\, t^2 E'(t) - \left[\int_{-\infty}^{\infty} dt\, t E'(t)\right]^2 \\ (\Delta\omega)^2 &= \int_{-\infty}^{\infty} d\omega\, \omega^2 \tilde{E}'(\omega) - \left[\int_{-\infty}^{\infty} d\omega\, \omega \tilde{E}'(\omega)\right]^2 . \end{aligned} \tag{16}$$

This problem is well known in optics and quantum mechanics where it goes by the name of the minimum uncertainty wave-packet[1] . The solution is a gaussian

$$E'(t) = \frac{1}{\sqrt{\pi\tau^2}} e^{-\frac{t^2}{\tau^2}} . \tag{17}$$

The Fourier transform of this gaussian is

$$\tilde{E}'(\omega) = e^{-\frac{\omega^2\tau^2}{4.}} \tag{18}$$

and the keying wave form with this $g(t)$ has $E(t) = 1/2[1 + \mathrm{erf}(t/\tau)]$

$$A(t) = \frac{1}{2}\left[\mathrm{erf}\left(\frac{t+T/2}{\tau}\right) - \mathrm{erf}\left(\frac{t-T/2}{\tau}\right)\right] \tag{19}$$

where $\mathrm{erf}(x)$ is the error function[2] defined to be

$$\mathrm{erf}(x) = \frac{2}{\sqrt{\pi}}\int_0^x du e^{-u^2}. \tag{20}$$

The differences in the keying wave form are shown by the turn on shape of the keying pulses in figure 5.

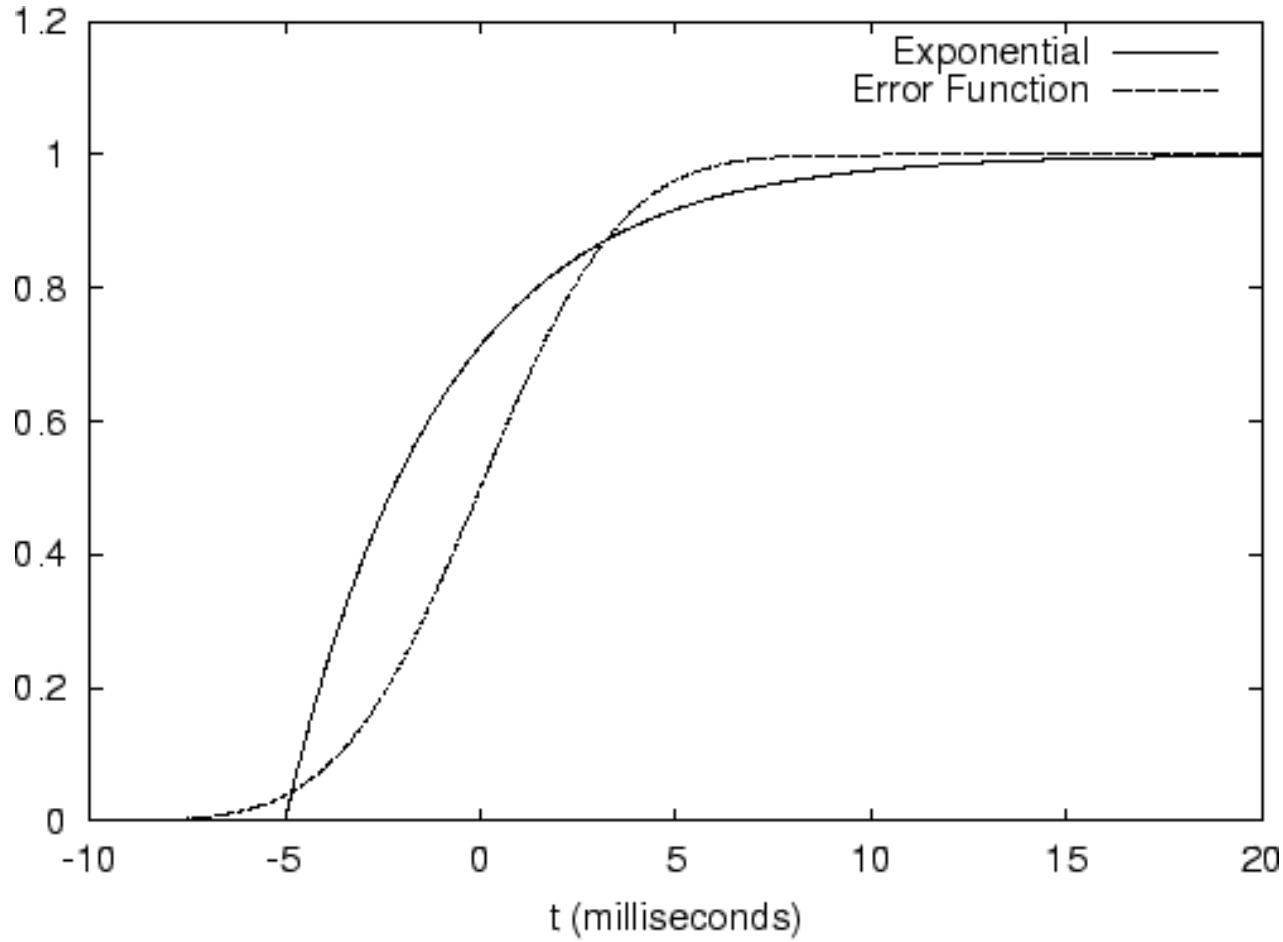

Figure 5: A comparison of the error function turn on with the exponential turn on, both for their respective τ values of 4 milliseconds.

In figure 6 I show the keying wave form for $\tau = 4$ milliseconds and $T = 20$ and 50 milliseconds as in figure 1 for both the exponential and optimized wave form. Notice that the abrupt changes in the exponential form are absent from the error function form.

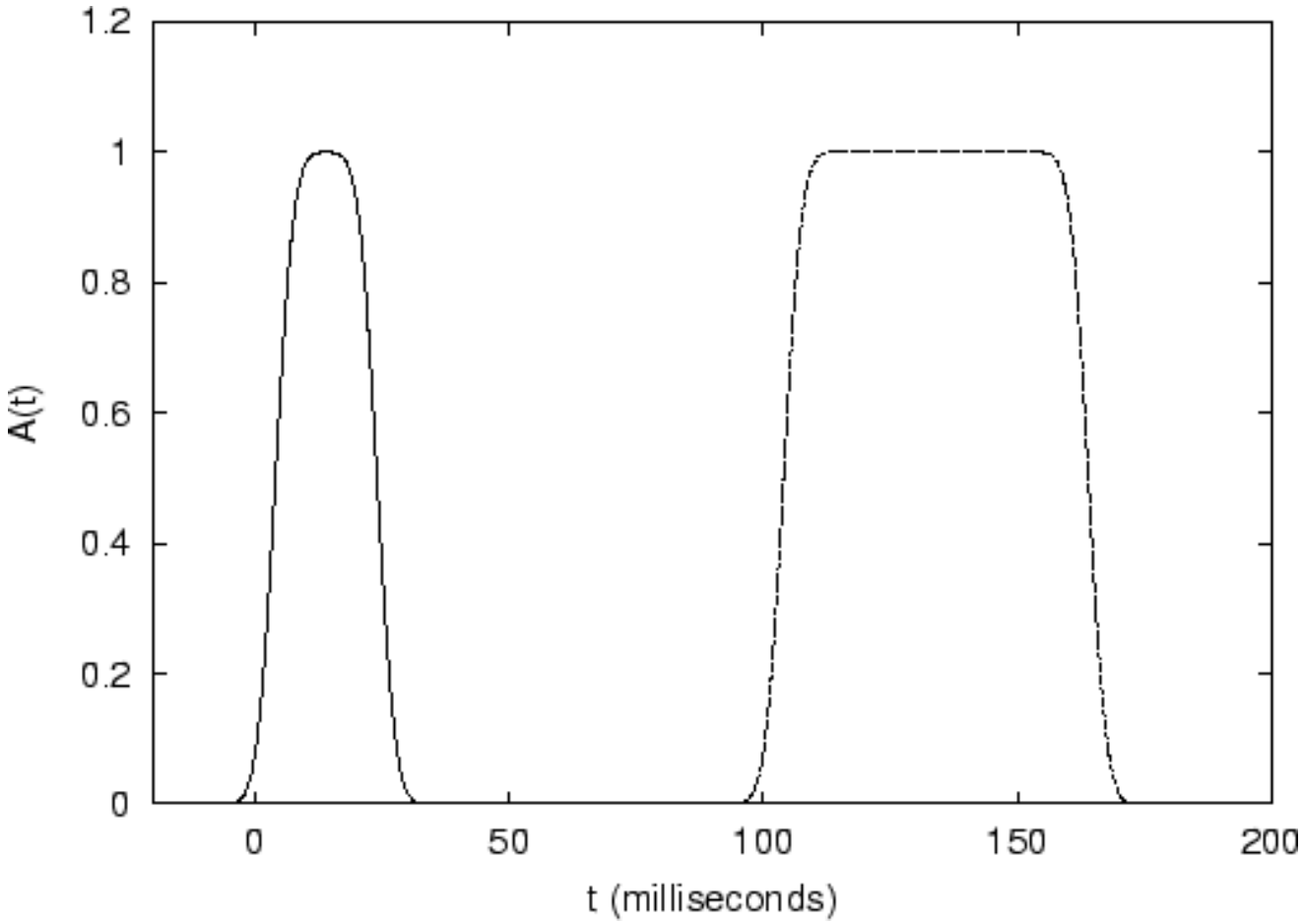

Figure 6: The error function keying waveforms for τ of 4 milliseconds and durations of 20 and 50 milliseconds. The 50 millisecond pulse begins at $t = 100$ milliseconds to separate it from the 20 millisecond pulse.

Figures 7 and 8 show the sideband energy density.

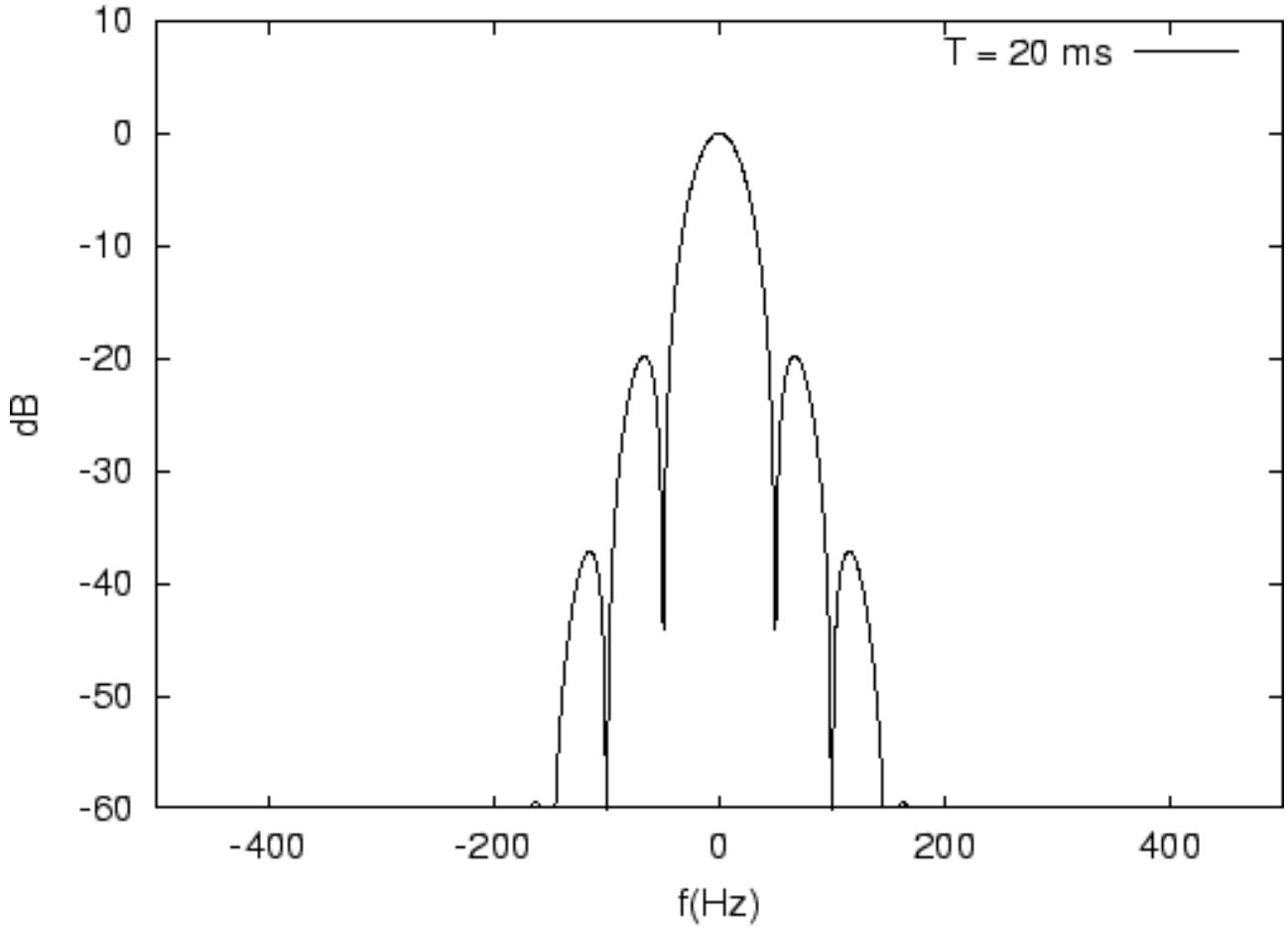

Figure 7: The sideband energy density for an error function keying wave form with $\tau = 4$ milliseconds and T of 20 milliseconds. The carrier energy density is set to 0 dB.

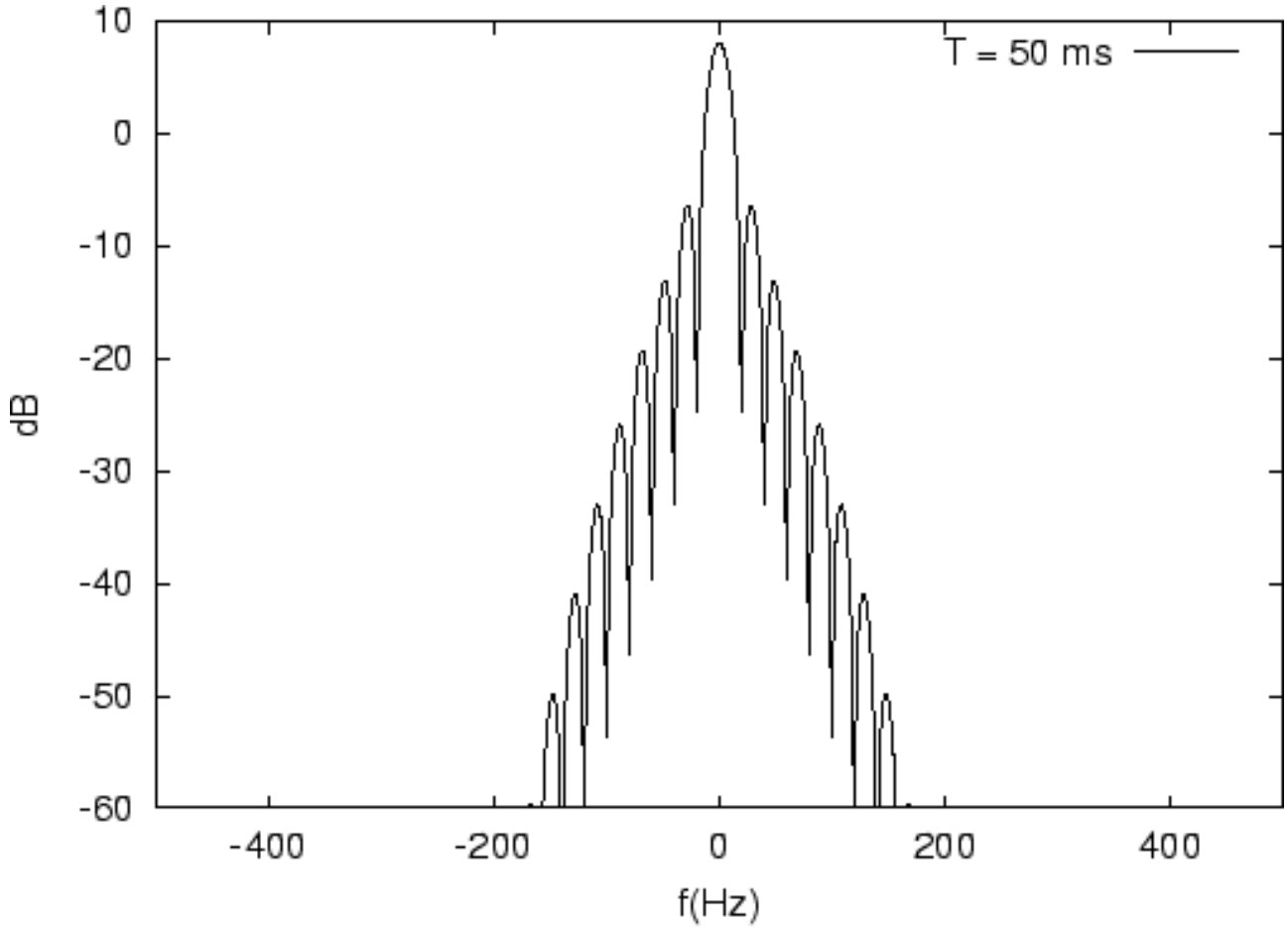

Figure 8: The sideband energy density for an error function keying wave form with $\tau = 4$ milliseconds and T of 50 milliseconds. The scale is the same as figure 7.

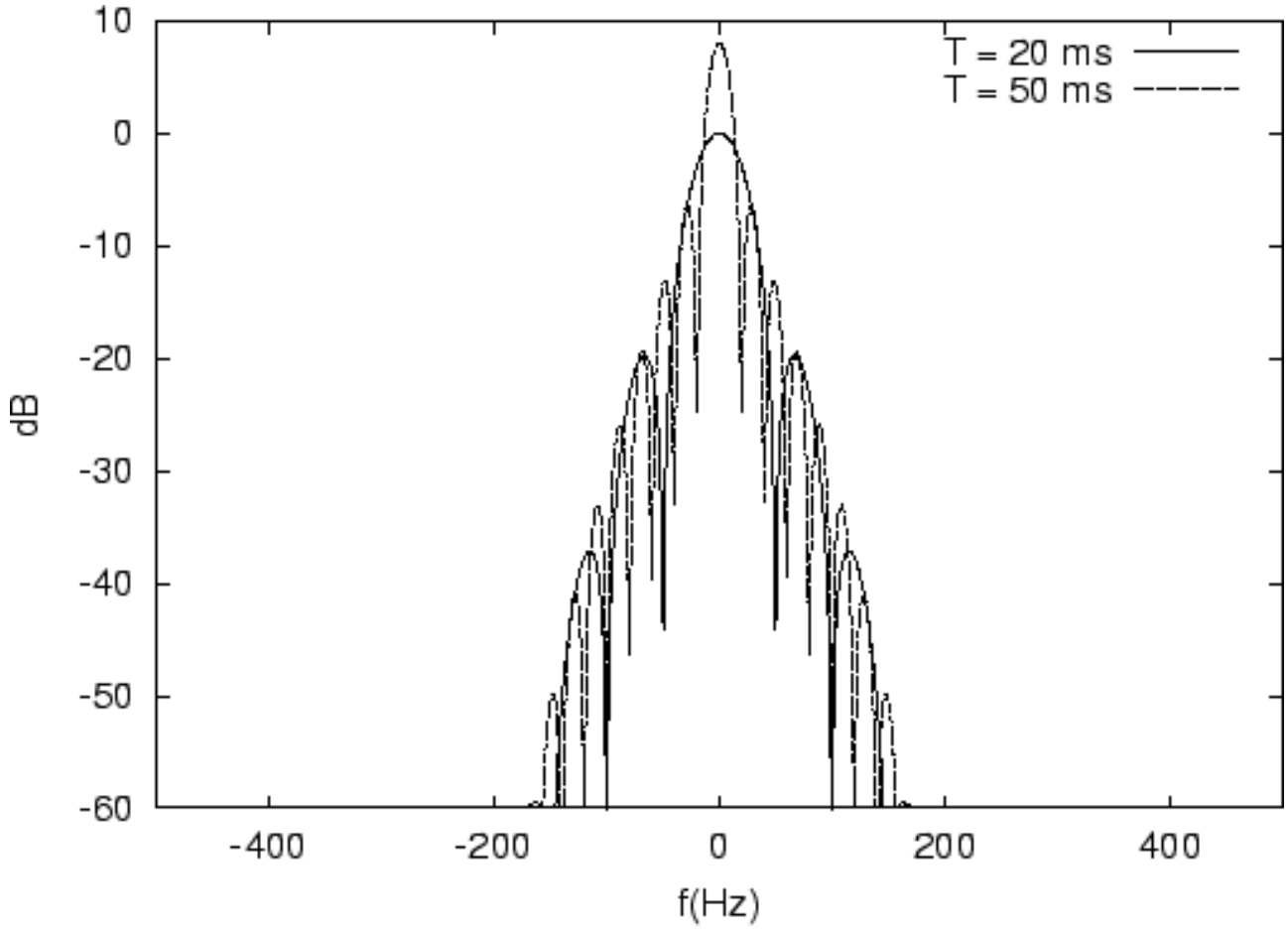

Figure 9: The plots of figures 7and 8combined.

5 Extensions and Conclusions

The sideband energy density for many pulse shapes factorizes. The first of the two factors is proportional to the Fourier transform squared of the square pulse and the second by the Fourier transform squared of the slope of the rise and fall. The analysis can be easily generalized to an arbitrary sequence of pulses. The Fourier transform of the single square pulse simply needs to be changed to the Fourier transform of the sequence of square pulses. For a given rise time, the error function shape for the rise and fall will attenuate unnecessary interference away from the carrier frequency much better than exponential keying.

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1

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2

M. Abramowitz and I. Stegun, *Handbook of Mathematical Functions* (National Bureau of Standards, New York, 1964).

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Up: W9CF Links

Welcome to the International Radio Web Site!

http://www.inrad.net

About key clicks

At International Radio we get a lot of questions about the FT-1000MP series key clicks and the effect of the Inrad mod. There is a lot of data out there, some of it taken with spectrum analyzers and some taken with second radios. There is no standard for comparison in the spectrum analyzer data, so it is difficult to draw any conclusions from it. Data taken with a second radio is very subjective at best and is essentially meaningless for comparison purposes.

We have taken data on unmodified FT-1000MPs (original version) and also on radios modified with the Inrad circuitry. There have been no changes made in the keying circuitry in the Mark V or the Field versions. The data was taken on the same spectrum analyzer, with the same settings and with the analyzer settings clearly shown on the plots. In order to compare the MP performance with some other radios, we also are showing data taken under the same conditions, with the analyzer set the same way, etc. on several other radios. We will post data on our site on any one radio of a type which we can get for testing.

Here is the technical explanation for the measurements. The radios are being keyed at 40 WPM by an external keyer. The spectrum generated by a CW transmitter is similar to amplitude modulation where the modulating frequency is the keying rate. Sidebands are produced on both sides of the carrier. If the modulating source were a perfect sinewave, there would be only one sideband on each side of the carrier. However, the modulating source is nearly a square wave and is very rich in harmonics. So the sidebands created are spaced by the modulating frequency and have an envelope which is a function of the waveshape of the keying source. Note that the envelope does not change for different keying speeds. The number of sidebands within the envelope change as the keying speed is changed, but this does not alter the bandwidth of the composite signal.

The really important thing here is the shape of the keying signal.

If we pick a certain offset from the carrier and take a dB down reading, we can have a figure of merit for comparison purposes. One kHz is a good place to make this measurement as the sideband level is generally way out of the noise

and is easy to read from the plots. These numbers are posted next to the analyzer plots.

Several things are readily apparent when viewing the plots. First, the Inrad modification makes a substantial reduction in the FT-1000MP spectrum. Also, the modified MP is quite good when compared with the other radios. As a matter of fact, it's narrower than some radios for which popular opinion claims good keying.

The "bad name" which the MP has developed in recent times probably has something to do with the number of radios in the field. In a major contest there are more MPs in use than any other radio. Imagine, for example, if everyone in the last 160 meter contest was using an Omni VI+ or a K2 driving a KW amplifier. The popular press would then be saying that something has to be done about those radios. In truth, something needs to be done with most of the currently popular radios.

The Inrad mod

What is it and how does it work? The mod consists of 3 small components, two go on the IF board and one goes on the RF board. On the IF board, a 0.047 capacitor and a 100 K resistor in parallel are placed across C2148. On the RF board, a 0.1 uF cap is placed across C1004 (Mark V and Field) or C1216 in the original MP. (If you don't have a service manual, Louis, W7DZN has posted pictures on the VA3CR web site.)

In each case, these capacitors are used to slow the waveform which is used to key the gate voltage on FETs. These stages are doing the CW keying. The resistor on the IF board is required to keep the duty cycle of the keyed waveform at 50% by rebiasing the gate of the FET.

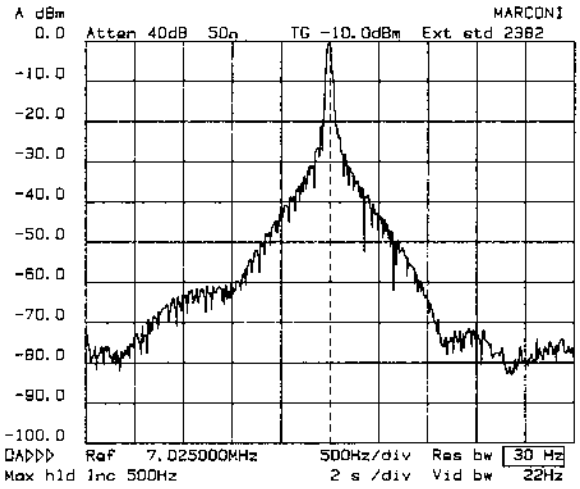
Some people have asked about routing the RF through the 500 Hz filter instead of the SSB filter. It seems like a good idea, however, the keying is done beyond the filters and this change would have no effect on the keyed waveform. Note that this is done in the Omni VI+ and the keying is distorted by the ALC loop. Another way to improve the keying would be to use a gausian or a transitional filter to shape the keying waveform. In a perfectly linear system this would work very well. However, the FETs being keyed are far from linear. Thus it is not possible to improve the output spectrum beyond what the Inrad mod does without substantial changes in the radio.

(continued)

2/4/03 W2VJN

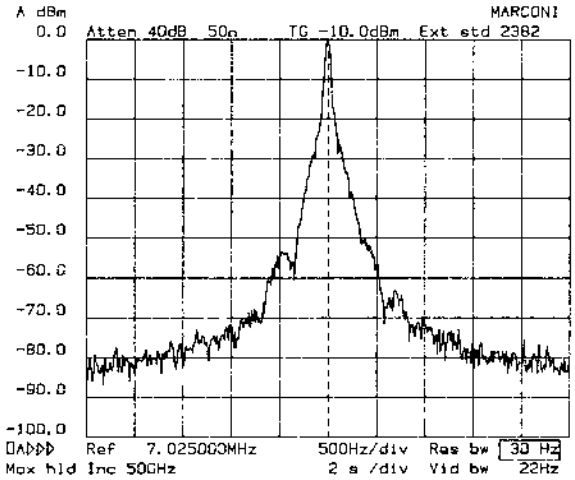
Unmodified MP

Level at +/-1 kHz= -62 dB



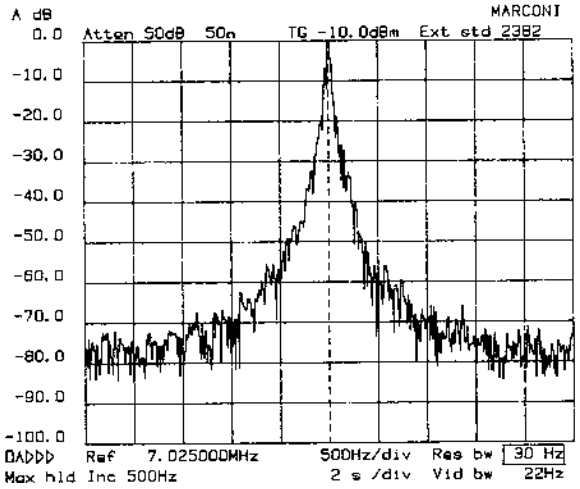
Modified MP-1

Level at +/-1 kHz= -73 dB



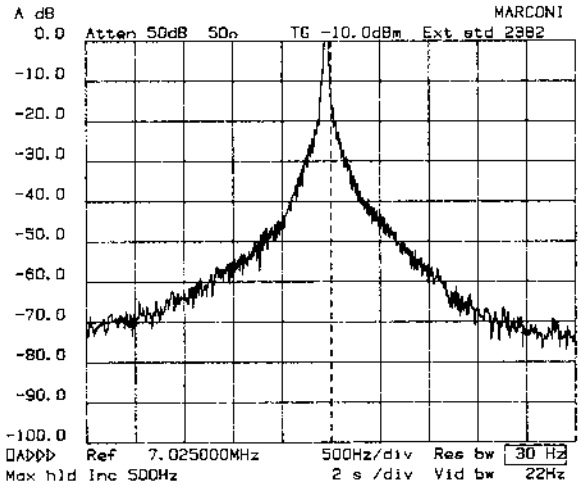
Modified MP-2

Level at +/-1 kHz= -72 dB



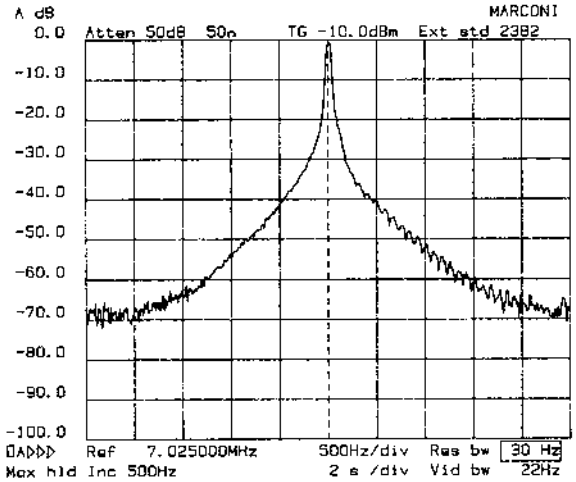
IC-765

Level at +/-1 kHz= -58 dB



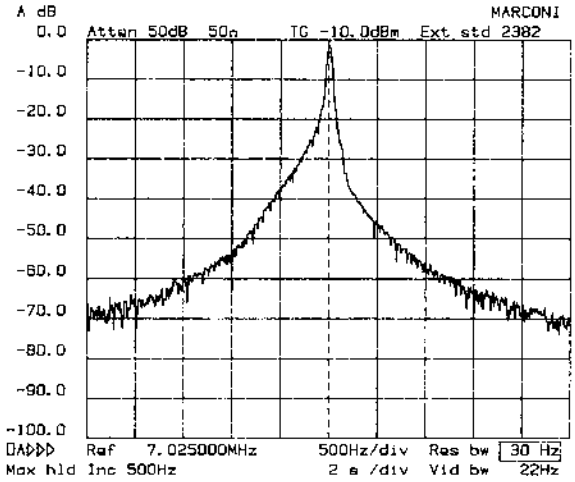
K2

Level at +/-1 kHz= -54 dB



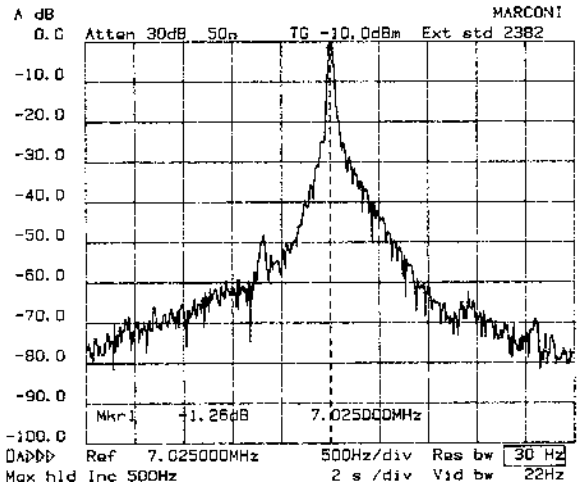
Omni VI+

Level at +/-1 kHz= -56 dB



FT-817

Level at +/-1 kHz= -61 dB



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• Beverage Antenna Products
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Software
Mobile Accessories
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Beverage Antenna Products

This equipment provides the best possible receiving system performance of any design! This is the same equipment design proven to make top scores in 160m contests year after year. Fully isolated feeds prevent common mode noise and unwanted signals. This equipment has excellent reliability and is housed in metal cases for improved life and shielding. Units use RF tight type F connectors for superior reliability and weather characteristics.

W8JI Beverage Antenna Construction Article

Reversable Beverage System

The RBS-1 is a two-direction, reversable Beverage array designed and used by W8JI. It consists of the reversable feed point matching system (RBS-1FP) and a reflection transformer (RBS-1RT). The unit can switch directions of reception by having a 10-18 Vdc voltage applied through the coaxial feed. The antenna between the two units is standard 450 Ohm ladder line (not included). Operating range is a broad 0.2-30 MHz. Fully isolated grounds are used to prevent common-mode noise and unwanted signals.

Specifications:
Power Requirement: 9-10 Vac or 12-15 Vdc
Switching Current: less than 80 mA
Antenna Impedance: 300-600 Ohms (450 Ohms recommended)
Feedline Impedance: 50-75 Ohms (75 Ohm recommended)
Reflection Transformer Ratio: 1:1

Metal housings are used for shielding and improved life.

View Assembly Instructions for DXE-RBS-1P (314K pdf)

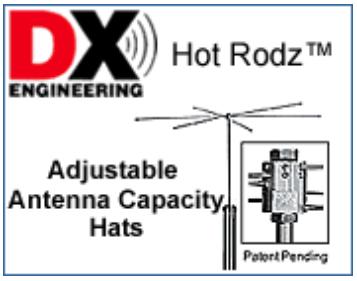
Right Click, “Save Target As”

Announcements

NOTICE:

DX Engineering will be CLOSED Monday, September 6 in observance of Labor Day.

New Products



Customer Showcase

Walton Gibson
K4LLA
Harvest, AL
USA

Get more information about this antenna >>

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Be an Antenna Superstar!
You put a lot of time and effort into designing and building your DX Engineering-equipped antenna. So why not show the world what you have created by submitting it to the DX Engineering Antenna Showcase? It’s easy to do, too. Just click on the Entry icon above, fill out the information form--don't forget to include information about the DX Engineering products used on your antenna, attach some photos, and presto!--you’re an antenna

Part #	Price	Add to Cart
DXE-RBS-1P	$129.00	

superstar!

Receiver Preamplifier, 0.3-35 MHz

This is the best HF low-noise amplifier available. The RPA-1 is optimized for 0.3-35 MHz operating range. It's push-pull amplifier design and robust components enable it to with stand high signal levels and operate when you need it most.

Gain: 16 dB, 0.3-35 MHz (+1.5-0 dB over this range)
Input Third Order Intercept: 23 dBm
Noise Figure: 3.5 dB
One dB Compression: +26 dBm (~0.4 W output)
500 Hz BW IM3 Dynamic Range: 110 dB or greater
Power Requirement: 10-18 Vdc @ 140 mA maximum

Suitable for indoor or outdoor installation with the option of being powered through the coaxial feed. Metal housing provide shielding and improve life. RCA and Type F connectors are used on the input and output of the RPA-1. The unit comes with a relay that automatically bypasses the amplifier when dc power is removed.

Part #	Price	Add to Cart
DXE-RPA-1	$109.00	

Beverage Feed System

This Beverage Feed System is designed and used by W8JI. This receive only antenna pulls signals out of the noise that most antennas don't even pick up. The BFS-1 has fully isolated feeds to prevent common mode noise. The transformer used in this receive only antenna handles 25 watts of power so it won't go into saturation even with the strongest of signals.

Specifications:
Operating Frequency: 0.1-30 MHz
Antenna Impedance: 400-500 Ohms
Feedline Impedance: 30-100 Ohms (75 Ohms recommended)
Common Mode Isolation: 20-30 dB (antenna to feedline)
Power Handling: 20 watts continuously or 200 watts for 5 seconds
Termination Resistance: 400-500 Ohms

Utilizes stainless steel hardware and metal housing for long life. Utilizes type F connectors for superior reliability and weather characteristics.

View Assembly Instructions for DXE-BFS-1(316K pdf)
Right Click, "Save Target As"

Part #	Price	Add to Cart
DXE-BFS-1	$35.00	

470-Ohm Non-Reactive 2 Watt Resistor
Pack of 10 Resistors

This non-inductive resistor can absorb extreme amounts of power for short periods without damage. It has the highest surge immunity of any 2-watt leaded (with leads) carbon or metal resistor. Metal-film and carbon-film resistors easily fail from even minor electrical disturbances. When used as a Beverage termination, the DXE resistor withstands nearby lightning strikes significantly better than hard-to-find carbon composition resistors. It is superior to carbon composition resistors in real life, retaining resistance value despite exposure to heat and power surges.

Applications:

- Beverage Antenna Terminations
- Parasitic Suppressors
- Dummy Loads

Sold Only as a Pack of 10 resistors

Part #	Price	Add to Cart
DXE-ECM-R470-2	$20.00	

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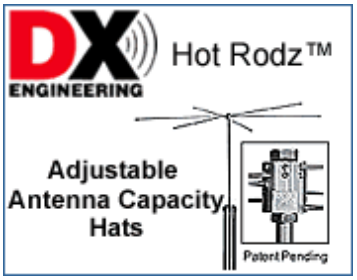

Customer Showcase

Anna Hill
KG8YL
MI
USA

Get more information about this antenna >>

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You put a lot of time and effort into designing and building your DX Engineering-equipped antenna. So why not show the world what you have created by submitting it to the DX Engineering Antenna Showcase? It's easy to do, too. Just click on the Entry icon above, fill out the information form--don't forget to include information about the DX Engineering products used on your antenna, attach some photos, and presto!--you're an antenna

superstar!

Downloadable Files

NOTE: I place no restrictions on the use of any of these files except that they may not be sold. I do not warrant the programs in any way, and no support is available for them.

All downloadable files with .EXE extensions are self-extracting ZIP files. After downloading a file, go to the directory containing the file (first starting DOS if you're using a Windows system). Then type the name of the file. (You don't need to type the '.EXE'.) This will cause the file to expand into two or more other files. One of these will be a documentation file such as READ.ME or INFO.TXT containing instructions. Windows users can generally read these files with Notepad. Files with extension .ZIP must be un-ZIPped with PKWare's PKUNZIP program, widely available as shareware. Again, a documentation file will appear when the file is un-ZIPped.

The "Field Day Special" Antenna

The Field Day Special is a two-element wire beam with a unidirectional pattern. It's similar to the "ZL" special, except that its elements are the same length. This makes it easy to reverse the pattern by using two feedlines, one at the center of each element. The original antenna was described in June 1984 QST. An antenna like this can't directly be scaled for other bands unless the size of the twinlead it's made from is scaled in proportion. However, using the program available here, you can design your own for the frequency band of interest. By changing the various dimensions, you can trade gain/beamwidth, feedpoint impedance, and bandwidth. Improvements in the program have been made since the article was published, so an optimal design for 20 meters will have dimensions slightly different than the ones shown in the article. The currents in the elements of one design were measured and found to agree very closely with those predicted by the program. Quite a number of these have been built by various people for several bands, with universal reports of operation as predicted. I use a 20 meter version nearly every year for Field Day, and occasionally 15 and 40 meter versions as well. 20 and 15 meter Field Day Specials have been my primary home QTH antennas for these bands for a number of years. Download **FDSP~.EXE**, then follow the directions under "Downloadable Files", above. More information about the antenna, including construction and feeding, is included in one of the expanded files. The program is a DOS program and can be run from the Windows 95/98/NT DOS prompt.

Current Baluns

Baluns are a continuing source of discussion on the Internet newsgroups. In response to some questions about "current" or "choke" baluns, I wrote an analysis of their operation. In the past, I had believed that placing the balun at the input of an unbalanced tuner would allow the balun to work properly when the output of the tuner sees a high impedance load. The analysis shows this to be incorrect. A brief analysis is given in **IBALBRF.TXT**, and a more detailed analysis in **IBALUN.TXT**. Both are plain ASCII text files and can be viewed with Notepad or a similar application.

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RF — Test and Measurement Homepage

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- 40 M Popcorn Superhet Receiver
- Broadband Transformer Topics
- Diplexer Topics
- Ugly Construction
- Wee Willy 75 M DSB Transceiver
- W7ZOI JFET Biasing Tutorial
- EMRFD Review
- Tapped Cap Impedance Transformation
- More Active Antenna Experiments
- Fun with LEDs
- SWL Receiving Antenna Experiments

Topics 2006 - 2009

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- TRF Receiver for WWV
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- Junk Box Low Pass NDB Filter
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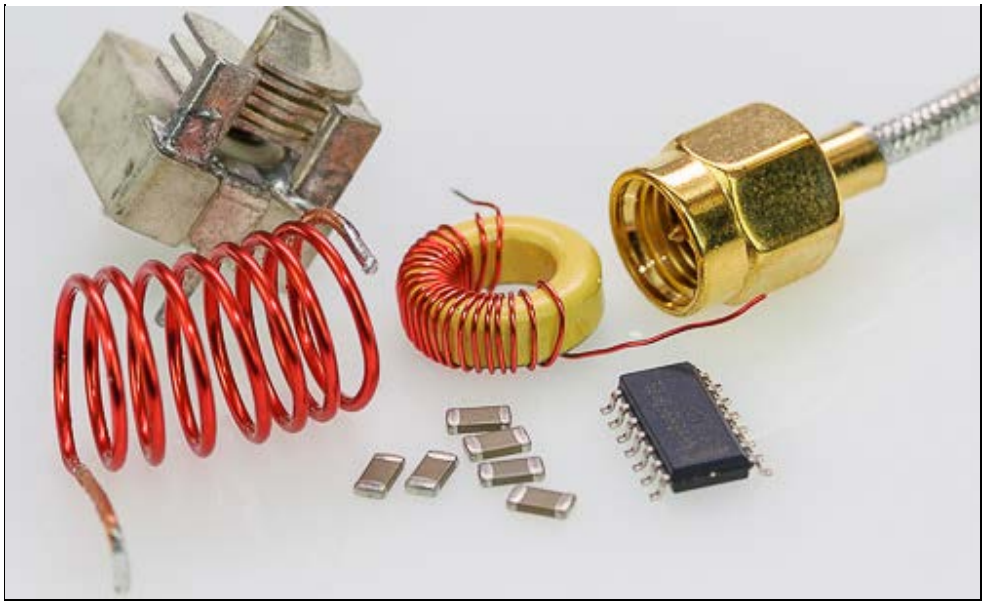


Make it — then measure it!

RF — Test and Measurement

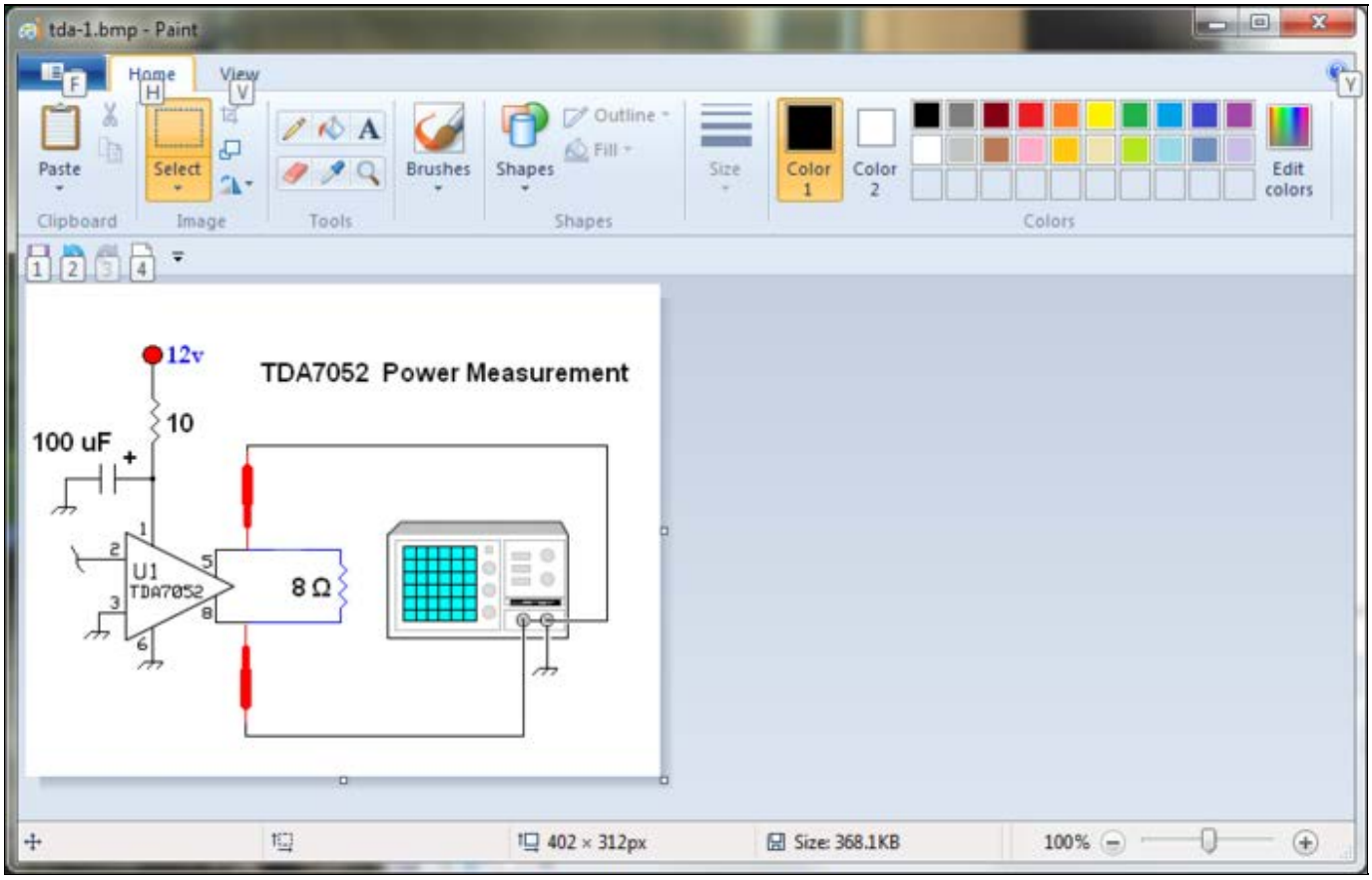
Frequently Asked Questions

1. How to you make your schematics?

I hand draw my schematics using the Paint Program that ships with all Microsoft Windows(tm) operating systems.

I strive to make schematics as clear and as small in file size as possible.The Windows Clipboard is used extensively to copy and paste the desired components from previous schematics to new schematics — very few new components are ever drawn; rather they are recycled from schematic to schematic. My system allows me to paste color 3-D labels and small photographic bitmaps into schematics.

The raw 24-bit bitmap drawings are compressed to 16-bit png files. Prior to May 17, 2010, bitmaps were compressed as 8-bit gif files. An example drafted circuit using the Windows 7 Paint program follows.



At least 25 people have sent or recommended software for making schematics. Thank you for this kind gesture, but I prefer my current method and after 15 years — make schematics quickly.

2. How come you don't supply parts lists? Other people do.

The answer is simple; lack of time. It takes considerable time and effort to put up a new web page and also to maintain a large web site. I save time by leaving the parts list up to the builder. In addition, this site is about experimentation and using what parts you have on hand.

3. Why didn't you answer my email?

I answer all legitimate emails as soon as possible. Our POP3 server gets an average of 2016 spam emails per month (December 2013 data), however, our software removes 99.27% of these and I never see them. Occasionally, legitimate emails are filtered in error and I apologize. I've received as many as 83 legitimate emails in a single day so can get behind. Please keep emailing — when readers stop emailing, I'll know the site has fallen totally obsolete [it may already be obsolete] and delete it.

Our mail server software logs and analyzes the network information of all spammers and may automatically filter and/or block their addresses or even their entire ISPs at the router level. Analysis indicates that 90% of our SPAM comes from just 3 countries and if you happen to live in one of these countries, the filtering will be especially sensitive.

This sounds dogmatic and unfriendly, however, until you've set up a domain and must handle ++ spam emails, endure and then develop router control software and other strategies to handle DDoS attacks and so forth, your completely naive about the 'dark side' of running a homebrew web server. Massive amounts of bandwidth might be otherwise wasted by allowing unwanted server use unless we actively counter these activities to keep the site running well for legitimate hobbyists. Not to mention all the wasted time. Further:

The 3 host web servers, the power, the server software, the security apps, the internet bandwidth etc.are owned or purchased by my family, and as you know, nothing is free. Despite many offers by companies to place ads on my pages, I've kept the site advertisment free and running pops.net costs us a few thousand dollars each year. We ask you to please respect our site for the sake of the experimenters who visit.

Pops.net server rack in our warehouse.

While I appreciate that some people might want to email invite me onto their social networks, I do not have time to participate. All email traffic from or involving social networks see this page for a list is deleted automatically by our POP3 server control software.

I **never** buy or sell parts via email, nor exchange hyperlinks. Never. I do give free parts to those in need though.

All email with the .info domain is blocked.

The number of people selling kits has jumped up by ~4 dB in the past 5 years. Increasingly, builders who need help with kits were emailing me for support. I rarely build kits and my knowledge regarding kit building is nearly 0. Please contact your kit seller for help.

You may wish to enquire with the kit seller about their online support polices and promptness prior to purchase. Additionally you might try the "support" email address provided and see if and how promptly they reply. Most of the popular kit sellers (AADE, Kits and Parts, etc.) provide excellent support to their customers. Like anything else online; buyer beware.

4. How come you didn't link to my web site - I linked to yours?

A big thanks to the folks who link to this web site! The QRP/SWL HomeBuilder site focus is content, not web links. Making a lot of links means spending time testing for and tracking down dead links - the so called "link rot". Time spent on the web site is time away from the electronics work bench. In addition, it is not logistically possible to reciprocate in kind, as hundreds of web sites and blogs have linked this site.

5. I see the word "popcorn" used a lot on this site- what's this all about?

Popcorn connotes the essential theme of the web site; simple, frugal, without fuss and over use of technical jargon, or complex math and engineering techniques. The QRP/SWL HomeBuilder web site is referred to as the popcorn site by many. The site targets hobbyists. The emphasis is fun. The hope is that it will attract new people to electronic design, measurement and experimentation. Hopefully, this site stimulates interest in QRP homebrew electronics.

Soon after I began building electronic circuits, my teachers and the popular electronic-related media of the day pushed me towards etched, printed circuit boards. I complied and this killed my passion for electronics. For me, habitually stuffing circuit boards lacks creativity and freedom.

Later, I discovered people were building guitar and bass amps using point to point wiring techniques with terminal strips and partial circuit boards. I became interested in building and repairing guitar amps and this passion continues today. In 1992, the discovery of 2 QST articles changed everything *for me (complete reference provided):* The Ugly Weekender: parts 1 and 2 by Roger Hayward, KA7EXM and Wes Hayward, W7ZOI; published in QST for August 1981 and June 1992. This was my first exposure to Ugly Construction and it was immediately adopted as the defacto standard bread boarding method in my electronics work shop. In fairness, etched circuit boards are a great tool, but not essential for the experimenter.
After working with Ugly Construction over time, considerable progress was made in understanding RF circuits and one output was the launch of this web site in 1998.

Currently, little has changed, I continue to prefer scratch-homebrew rather than kit-homebrew electronics. My interest in Short Wave radio and analog electronics has grown considerably. For me, electronic circuits hold a certain mystique which arouses my curiosity to learn, enjoy and share. As a lay person, this web site has facilitated meeting some awesome people through email from all continents and it has been a privilege to learn from them, my mentors, book and web authors and often enough; from my mistakes.

5. What do you mean by a 5K1 or 3K3 resistor value?

For E24 or 5% tolerance resistors 5K1 = 5.1K, 3K3 = 3.3K and so on. For E96 or 1% resistors 31.6K is written as 31.6K. All resistors are 1/4 watt unless otherwise specified.

6. How do you measure audio amp output power?

Please see Figure 4 on this web page. Any amp when cranked, outputs much greater power than when it is providing a clean sine wave. The quoted power for any audio power amp on this web site is the maximum average power it will give before the pure sine wave becomes distorted.

7. I noticed a new web page appears and then it is edited for 1-2 weeks. When is the web page completed?

When a new web page is added, it takes a week or so to find and change some of the grammar and spelling errors. Sometimes new ideas or feedback will cause me to further edit a web page at any point in time. This whole web site is a work in progress. The last date any given web page was edited is posted on the bottom of the web page.

8. Do you buy or sell stuff?

No and no. I receive numerous emails from people asking me to sell them stuff. I do not sell anything - no parts, books, coffee cups, ball caps, tee-shirts, ad space — nothing. I do not buy parts in commercial-quantity volumes and have no need to make contracts for obtaining any electronic components. Every week, Asian companies email to ask about buying their parts — please note, my answer is *always* the same: no thank you.

9. Questions and concerns about printing and printability

Each year, a few readers email to complain how poorly the web pages print. This is true and I apologize.

Some people prefer pdf files for easy printing. I have resisted going to pdf format for 3 main reasons:

1. The web site audience is international and many are using web translators. PDF files are 8-bit graphic image files and do not translate.
2. More and more readers are using mobile computer devices and pdf files are a pain for them.
3. We should all print less often to save resources

As an experimenter, I dislike crammed, small-size schematics and feel they should be drawn for *maximum* clarity. Therefore, my schematics tend to have a lot of white space and color contrast. I try to make them no wider than 700 pixels, although sometimes it's impossible to do this. Big schematics are *not* printer friendly. The only practical solution is to click on and open them in a separate browser window for easier sizing and printing.

I also feature big photos which burn up a lot of printer paper. Project photos are important to me; they provide a more intimate glimpse into the bench work and promote the real purpose of the site — building stuff.

A potential printing solution for Microsoft Explorer 8 users; Click

10. I have noticed in your CMOS logic photographs, you don't always ground unused input gates. Isn't this bad?

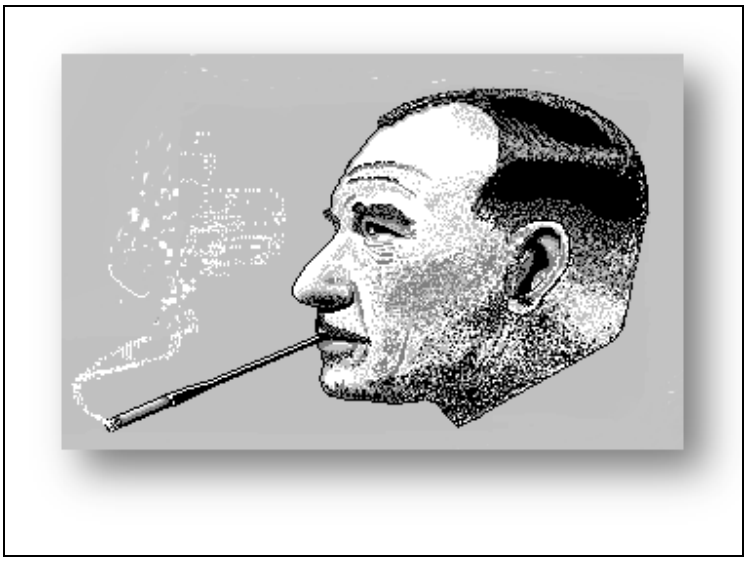
Proper CMOS logic practice mandates the grounding of all unused input gates. In prototypes and experiments, I don't always do this as I generally want to re-use the IC in other experiments. This is a cost saving measure. When you build a lot of stuff, it can get expensive and recycling parts makes sense. In keeper circuits or critical prototypes, unused input gates are directly soldered to the copper clad board. This also anchors the IC very well.

11. What is the proper URL of the home page?

http://www.qrp.pops.net

The following pre-2006 URL was decommissioned August 6, 2010: http://www.qrp.pops.net/default.htm

12. What are the QRPHB Design Centers and Professor Ivanenko character about?



This web site is about design and not just providing circuits to copy; I'm hopeful that the QRPHB Design Center concept initiated in 2011 will invigorate the site. Design Centers are the presentation of simple, but useful algorithms for amateur builders to advance their skills on the bench.

Professor Vasily Ivanenko ( ), a *fictitious* retired Russian physics professor wants to share his knowledge and give back to society. He signifies each Design Center. Professor Ivanenko was drawn for me by Rod Adams in 1996 using the Paint program that ships with Windows (the same app I make my schematics with). Rod did all of the other original bitmap art for this website including the coil guy and junk box pictures. This character was inspired by one of my favorite photographers: Irving Penn — this photograph, which is all over the web.

A new character; Dr. Natasha Petrovna appeared in late Summer 2012.

The professors are just a good bit of fun — add intrigue, characters on whom to focus and a means of identifying Design Centers. Electronics with just math and physics bores us all. Adding splash, color, clear photographs and characters such as the coil guy or the Professors boosts the site's appeal and provides a creative outlet for me.

13. Why did you kill your blog?

Time mostly. *My* blog wasted *yours* and my time. I carefully analyzed my personal yield from blogs in 2012 to 2013 — for the most part, blogs just entertain + share trivia, or rehash someone elses idea(s), or 'innocently' attract you in hope to sell stuff — and sometimes, just fulfill the author's need for attention. I don't seek, nor have time for entertainment or spectacle within my RF hobby and I certainly don't wish to waste *your* time. Each to his own, I suppose.

My analysis showed that unfortunately, blogs rarely boosted my understanding of electronic design or measurement practices. My ardent focus is to learn + improve and then pay some of this knowledge forward on a web site. Of course, everything in context — many exceptional people blog. For example, Dave AA7EE, or Jason, NT7S.

Most of the RF design and measurement people I follow keep old fashion web sites and provide generous email support. Design and measurement web sites, plus reputable and/or peer-reviewed industry and hobbist books, journals and multimedia work best for me.

Further, great elmers don't just publicly hang-out on blogs, or web server groups, or publish Utube videos — some just check their emails and when asked — give wonderful support without fanfare. Hats off to these humble folks. Thank you!

14. What oscilloscope should I buy?

Yikes — a tough question I get nearly every month. Please do your research. My best answer is buy the best 'scope you can afford. Are you a casual experimenter, or sit in your lab a lot? I prefer DSO's , however, made due with a old boat- anchor CRT for my first 10 years. The Rigol 1052/1152 seem popular entry-level choices due to their cost versus performance ratio. On the other hand, view this video to see how much Rigol DSO technology has changed.

I owned and sold my 1052 to a builder in Michigan — a worthy choice like many other 'scopes. IMHO, the FFT and math functions on the 1052 and 1152 suffer due to low memory depth and clock jitter.

Even if you only work at HF, a bandwidth 5X higher than your main frequencies of interest works better for showing harmonics. Again, I advise people to simply buy the best 'scope they can afford since it will form the heart of your test bench.

15. I'm a beginner — what toroids should I buy?

Opinions will vary, but here's what I recommend. Buy any quantity you wish, but sometimes minimum quantities apply and shipping to some countries costs dearly, so I tend to order enough parts to last me for awhile.

A basic HF toroid starter kit might look like:

Quantity	Part#	Type
10-25	FT37-43	ferrite
10-25	T50-6	powdered iron

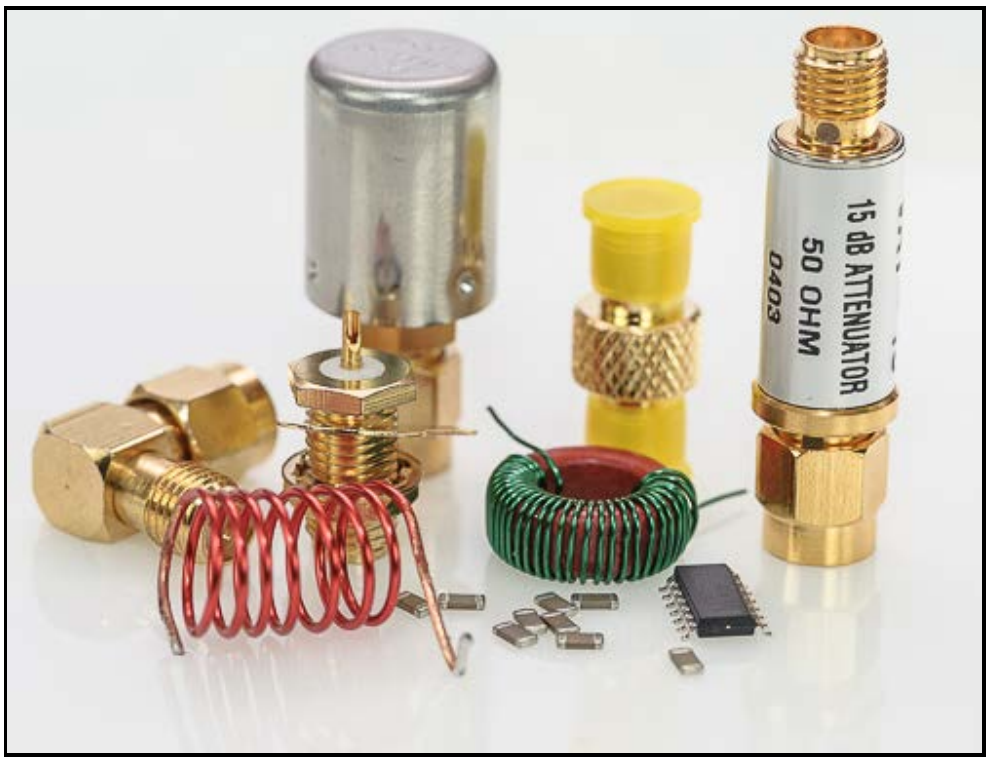
Above that, perhaps add these toroids.....

10-25	T68-6	powdered iron
10-25	T37-2	powdered iron
10-25	T50-2	powdered iron
10-25	FT50-61	ferrite

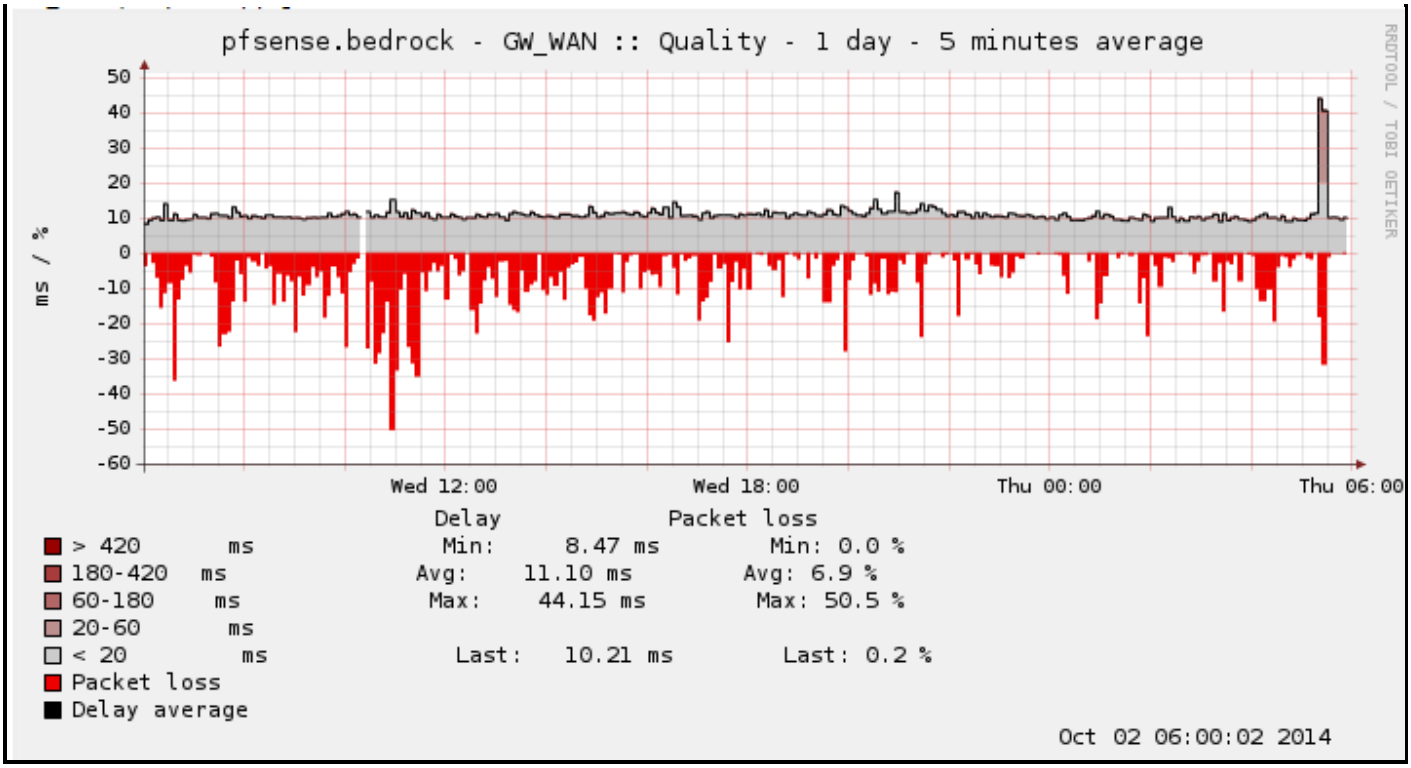
That's about it -- These will build most things RF on a HF beginner bench

RF — Test and Measurement

QRP — Log: Updates To Permanent Content



Oct 5, 2014 --- The end of QRPHB? Expect the Qrp site to go down each day for 1-4 hours. As I wrote before, we're losing an average of 10% of the outgoing packets and this peaks as high as 40% when many people are accessing the site. See an example graph below.



On Oct 2, 1 of our ISP's techs came and tested our WAN -- "it's better than most" he said. Hopefully they can do something to boost the performance on their side. Wave bought out all the local competition and we're stuck with them [this is happening everywhere]. Click for 1 of the many links writing about the emerging internet cable company monopoly.

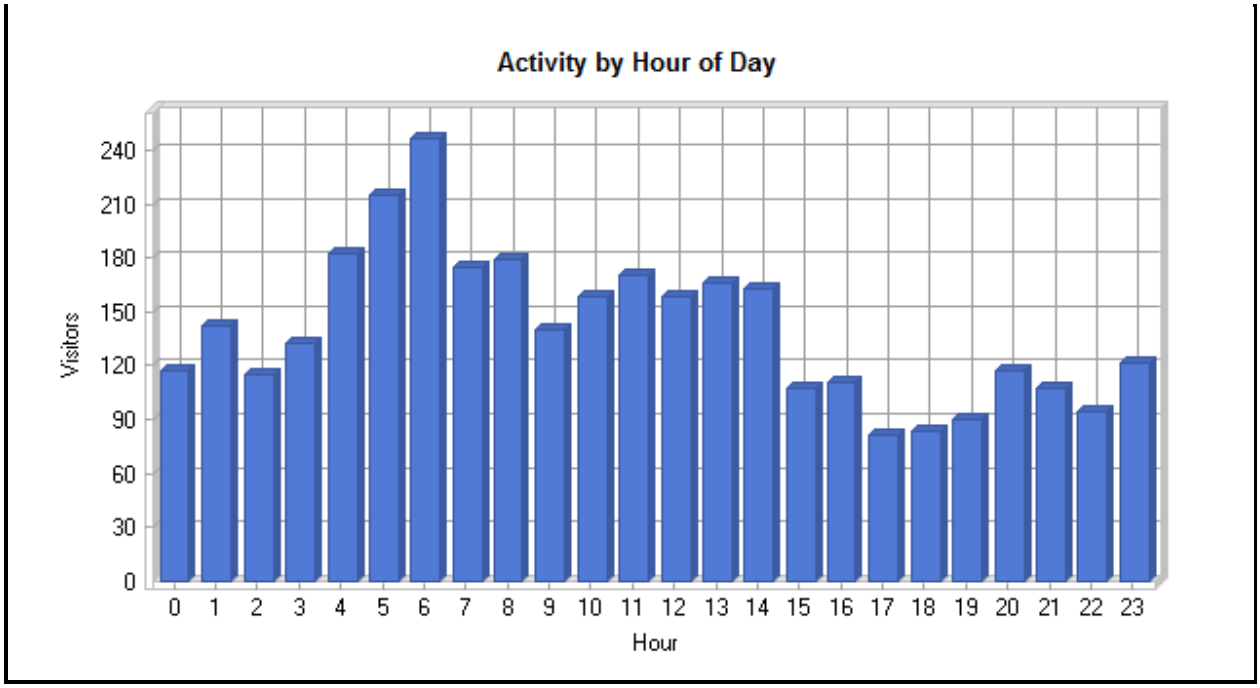
You'd think that in 2014 we could properly host a web site like we've done since 1998? To boot, Stuart is paying for a high bandwidth business account and getting less than 1% of the promised bandwidth. We chatted today and will wait patiently for Wave to improve our outflow - but only for so long.

If Wave can't or won't boost performance, I am leaning towards taking this web site down as opposed to moving it -- it's been a good run. QRPHB creates a lot of work and I could get way more done without the web site hassles. So, 2014 might be it for me. If so, thanks for coming here all these years and best of luck with your experiments!

--- Sep. 19 --- Wave Broadband, our Internet provider suffered major problems recently and I shut the site down for 4 days rather than have it limp along pathetically. Well the Wave saga *continues*: the site loads slowly and pictures are missing etc.

Our email server remains working even when the main site is down. You can always email me if the site is not working for you -- or if you have a comment or question.

Vistor volume rose this Summer so we added a better router that provides bit by bit data performance collection and new features to help throughput. At minimum user bandwidth, a unique visitor hits the site every ~0.7 minutes. I'll start formal bench experiments on October 3 --- to kick off the sites' 16th season.



August 31, 2014 — Kit to upgrade my HP frequency counter to 3 GHz added as Section 5 on Caitlyn 310 — UHF Beginnings.

August 18, 2014 — QRP-POS Data on the Sundry Web Page. Look at the end of Section 8: Popcorn AF Amps For Receivers — Reprise. This new, all-discrete AF stage will go into the Funster Line receiver and ranks as 1 of my best in terms of power and headroom.

August 12, 2014 — Funster Line: a QRP 40M band CW trans-receiver added to HF Embarcadero web page as Menu item 4. Click here for Funster. Only the transmitter is presented for now.

June 18, 2014 — I tweaked the page now called About... on the top level menu. Also, on this web page, I added a new essay for 2014 just under the essay for 2010. I'm off the bench until Fall, but hope to add a little content on rainy days or such. I'm about 2 years behind in presenting some of my experiments.

May 22 and 26, 2014 — New Supplemental Web Page for VHF-FM launched. This page and another supplement will house some new receivers over time. The new supplement is linked on the original VHF-FM web page in section number 4.

May 5, 2014 — Section 7: NE612 Mixer Diddy added to VHF Veronica

April 14, 2014 — Completion of Return Loss Bridge Experiments [added Bridge #4] on Caitlyn 310.

April 7, 2014 — Return loss and VCO experiments added to Caitlyn 310.

March 7, 2014 — Section 3 added to VHF-FM. A DC Converter for VCOs. **March 25** QRP-POSDATA for March 2014 Poor Hams Scalar Network Analyzer (PHSNA) added to Section 1 of Sundry.

February 14, 2014 — Caitlyn 310 — New repository web page for my venture into UHF. Click. Surprisingly, the site averages ~ 3000 unique visitors every 24 hours. Click for the Feb 25 tally

January 15, 2014 — The FAQ was often missed and therefore moved to the top level menu + editted/augmented.

January 4, 2014 — A follow-on version of the K3NHI QEX power meter added to RF Workbench 5 as section 6.

December 15, 2013 — I added Section 3 to HF Embarcadero. VXO and VCXO Notes. I significantly updated the VFO-2011 web page on Dec 17, 2013.

November 7, 2013 — Section 6 added to VHF — Véronique. 1 photo added to the end of the Ugly Construction page. Severe Fall weather

conditions took down the server for ~1 hour today.

November 1, 2013 — I added Section 2 to HF Embarcadero. A vestigal set of notes regarding my attempts to update the Popcorn Superhet receiver.

October 31, 2013 — Section 5 added to VHF — Véronique.

October 27, 2013 — VHF-FM re-formatted and Section 2 added. I imagine this page will disinterest many.

October 18, 2013 — 3 rarely accessed web pages removed from drop down menu, but not deleted from the server. New QRP-Posdata added to HF-Ragbag (near the end). Section 3 edited and Section 4 added to VHF — Véronique.

October 11, 2013 — Build Season 15 begins [I bench experiment Oct to May]. I added HF Embarcadero to hold all my HF and perhaps AF experiments this season. I'll also add content to VHF FM, VHF — Véronique and RF Workbench 6 over the next 12 months. Thanks.

September 9, 2013 — Blog deleted.

August 8, 2013 — Section 2 of RF Workbench 6 added. Measuring PA collector V and I to calculate efficiency. On August 16 — I added a new essay on the Ugly Construction page called Is Ugly Construction Less Reproducible than Manhattan?

June 25, 2013 — I started RF Workbench 6. It will take 1 year to complete

April 14, 2013 — I added Section 3: 50 Ω MMIC Bench Amplifer to the VHF 2013 Veronica web page.

March 31, 2013 — I started a new content page called Pin Outs

March 25, 2013 — Section 9: an essay about the 1981 Progressive Receiver, plus the final section — 10: Miscellaneous Pictures and Figures added to Sundry 2012-13

March 18, 2013 — Section 8. Popcorn AF Amplifier for Receivers — Reprise added to Sundry 2012-13

March 10, 2013 — I added a new VHF content page for 2013: VHF-2013 - Veronica

February 16, 2012 — I added Section **7** -- A Journey Above HF -- to the Sundry Experiments 2012-13 web page.

February 14, 2012 — I added Section **6** -- Non-Mechanical Iambic Paddle -- to the Sundry 2012-13 web page.

February 1, 2013 — A seperate QRP—POSDATA added to RF Workbench 4. Now QRP— POSDATA 1, 2 and 3.

December 15, 2012 — I added Section 4 to Sundry Experiments 2012 - 2013 . A PLL circuit from EMRFD.

January 5, 2013 — I added Section 5 to Sundry Experiments 2012 - 2013 . A simple AF feedback amp. An essay concerning L-C meters was also added on RF Workbench 5. Section **5.**

December 22, 2012 — I deleted the web page ***Tuning VFOs with a PN Junction*** since some of the experiments were poor quality and performed back in 1998 when I was more ignorant than now. I've learned much since then and my new VCOs from the past 1-2 years reflect this knowledge. The ***Selected QRP Reading list*** and ***Cascode 7 Receiver*** web pages were also wiped.

December 15, 2012 — I added Section 4 to Sundry Experiments 2012 - 2013 . A PLL circuit from EMRFD.

December 3, 2012 — I added Section 3 to Sundry Experiments 2012 - 2013 . Interview with Jason from Etherkit

November 13, 2012 — QRP — Posdata added to the VHF to the Max web page. Section 5: Z-Comm VCO.

November 1, 2012 — Sundry Experiments 2012 - 2013 web page added.
QRP — PosData added to Section 2 of **Power Meter Calibrators** on RF Workbench 5 and updated again on Nov 22, 2012.

October 16, 2012 — VHF FM web page added. Already, it has spawned a first supplemental web page

Sept 24, 2012 — QRP — Posdata added to the end of the HF Ragbag web page. I added a bypass and decouple network for HF to lower VHF.

August 17, 2012 — QRP — Posdata #2 added to bottom of the Receiver Band-pass Filters web page.

August 6, 2012 — Section 5: **Some Experiments with RF Bypass Capacitors** added to the HF Ragbag web page. Also a new QRP — Posdata added to the bottom of the Crystal Parameter Checker web page.

August 1, 2012 — QRP — Posdata added to the Receiver Band-pass Filters web page.

July 12, 2012 — I added a corrected schematic on the Wee Willy page: Wayne. M0WAY — 14 MHz PA under August 25, 2011. Also, added a new essay on **Microphonics in DC Receivers**. See Section 4 on the HF Ragbag Page

June 23, 2012 — RF Workbench 5 added. Click here.

May 31, 2012 — Galina discovered that I neglected to publish the proper version of the Hobby and Fun 2011 page and corrected my error. A fine-tuneable Wien Bridge Oscillator idea from Ken Kuhn now appears at the page bottom.

April 21, 2012 — I heavily edited RF Workbench 1 and 2.

April 13, 2012 — Web site purge. I removed RF Filters, VFO 1998, QRP Workshop Ideas, Miscellaneous Schematics and Photos, Base-biased VFO, Funster Transceiver, Miscellaneous Circuits and Ideas 2005 and Crystal Oscillator Offsets. **Reason:** substandard.

April 6, 2012 — 50 MHz Receiver Pre-amp and Filter added to VHF to the Max web page .Section 4.

March 26, 2012 — HF Ragbag web page added to top-level menu. Non-VHF experiment repository for 2012.

March 19, 2012 — VHF to the Max web page added to top-level menu. 50 MHz VCO experiments added to this page in Section 3.

March 14, 2012 — EMRFD review edited.

February 29, 2012 — Minor edits to the Audio Transistor Input Impedance Experiments web page. Also, I updated the calculation of the common emiiter amplifier base input resistance using the better formula: **Rin = (B+1)*(re + RE')** [while ignoring REB]. This is my favorite web page on the site.

February 26, 2012 — QRP— Posdata added to the bottom of the 2nd NDB web page

February 17, 2012 — I introduced a new miscellaneous VHF page: VHF to the Max — I'll slowly add stuff over 2012. Major editing done to the Broadband Transformer web page.

February 4, 2012 — RF Workbench 1 and 2 significantly edited. 2 new photos added.

January 28, 2012 — QRP — Posdata added to Crystal Parameter Checker web page and to QRP Modules 2011 under 7 MHz VCO Experiments on this web page. I re-wrote the temperature compensation section of the VFO-2011 web page and added 3 photographs.

January 3, 2012 — Web site change: I update web essays with an "epilogue section". In 2012 and on, they will be called QRP — Posdata (Spanish for post-script or epilogue). Posdata #2 added to the RF Workbench 4 web page.

December 17, 2011 — The Butler Did It ! - First VHF Experiments 2011 web page added. A 50 MHz frequency doubler added Dec 26, 2011.

Nov 12, 2011 — Our server went down for 16 hours. Both AC power and cable Internet to the warehouse failed after a rain storm and wind gusts knocked down some trees that severed the hydro and cable wires. Expect more weather-related down time as Winter approaches.

Oct 15, 2011 — VFO-2011 added.

Oct 2, 2011 — RF Workbench Page 4 added. On **Oct 17, 2011** I added an epilogue.

Sept 20, 2011 — Double Stacked Toroid VFO 2008 web page pulled off. It was substandard and some of the material will re-emerge on a VFO 2011 page this Winter.

Sept 19, 2011 — I updated the SWL essay since it was 6 years old and much has changed with respect to Internet radio. Over time I have received ++ emails expressing different views. I am more a SWL than a Ham and offer just 1 opinion and live by a "each to his own" mantra. Context is everything - this is a radio electronic experimenters site that recognizes SWL'ers are important members of the radio community. Click. Wee Willy web page updated again!

Sept 12, 2011 — Design Center concerning popcorn receiver band-pass filters added. Click.

Aug 26, 2011 — Minor update at the end of Wee Willy DSB transceiver. ***New Junk Box Blog format***. Change is good.

July 30, 2011 — New web page QRP Modules listed on the main menu. Currently under construction.

March 31, 2011 — 2 photos added to the Ugly Construction web page

March 19, 2011 — New content; Hobby and Fun 2011 . I'll slowly add more stuff over the year.

Feb 12, 2011 — New content. Miscellaneous RF Experiments 2011

Dec 29, 2010 — New content. RF Workbench Page 3

Dec 12, 2010 — Final additions to the the 2010 Hobbyist Page added — these concern matching FETs, BJTS and diodes.

Nov 10, 2010 — Some editing and 2 photos added to RF Workbench page 2.

Oct 9, 2010 — This October marks the 12th season of experiments for the site. I have 4 partially completed web pages on the go — pure craziness. I decided to finish them 1 at a time and then add them sequentially. Today, a new web page was added and is 1 of 2 supplements to a future main QRP audio page: Audio Transistor Input Impedance Experiments. The first new "permanent" content since March 2010.

Oct 7, 2010 —The Junk box page lay out was simplified: Accommodating the various modern Web devices plus screen resolutions proved difficult with the old html code. Although less impressive, the new format updates quickly and looks the same on every computer.

Aug 15, 2010 — The cable supplying the Internet connection failed. The site went down for 25 hours. Expect more shut downs over the next couple of days as we resolve any remaining problems.

August 6, 2010 — I decommissioned the historic home page URL http://www.qrp.pops.net/default.htm. This page was a html hard-coded parallel version of the the *correct* home page URL http://www.qrp.pops.net. It was just too much work to continue to update the old (pre-2006) home page in addition to the proper home page.

June 28, 2010 — Web subtitle change : **Amateur Radio Electronic Design** to shorten the name and reflect the site's main purpose.

Effective May 17, 2010 — Schematics now use the 16-bit png format. I have abandoned the 8-bit gif format

May and June 2010: The pops.net net control crew rest up for the next big wind storm. The servers lost power for many hours on May 4 and June 12 due to bad weather. Severe storms arising from the Pacific Ocean threaten our AC power lines each Spring.

March 2010

Wide Range L-C Oscillator added to Hobbyist 2010 page. Link **March 7**

JavaScript Applet **K** added to the QRP Tools page **March 11**

RF Workbench Page 2 added. Link **March 15**

February 2010

JavaScript Applet **J** added to the QRP Tools page **Feb 6**

Editing of the Low Noise Crystal Oscillator web page **Feb 22**

RF Workbench Page 1 added. Link **Feb 18**

January 2010

Experiment #6 added to the Hobbyist page **Jan 2**

New content Low Noise Crystal Oscillator **Jan 10**

2 photos added to the bottom of the Ugly Construction web page. **Jan 12**

1 photo added to the Broadband Transformers page **Jan 12**

New content Hobbyist Page 2010 **Jan 13**

New content: Supplement to JavaScript Applet G which is located on the QRP Tools page **Jan 11**

Additions to the Low Noise Crystal Osc page: the DSO versus CRO essay, plus 5 MHz crystal oscillator added **Jan 31**

JavaScript Applets H and I added to the QRP Tools web page **Jan 31**

December 2009

It has been brought to my attention that the email replies I am sending are not compatible with some of the latest email software such as Thunderbird etc. The reason was that I was using a homebrew email program written over 10 years ago. It is now obsolete. A new email platform is now in place along with a completely **new email address**. Consult the email web page for more information. **Dec 30**

Experiment #5 added to the Hobbyist page FAQ updated and edited. **Dec 30**

Experiment #4 added to the Hobbyist page **Dec 26**

Experiment #3 added to the Hobbyist page **Dec 23**

What does the output of a diode ring mixer look like in your oscilloscope? This has become a FAQ. The question is answered in a contribution by Wes, W7ZOI. Big thanks to Wes for this content. **Dec 22**

I am very pleased to present the **Mike, KL7R Memorial Receiver Experiments**. Click here **Dec 19**

New bulleted list format added to this page to improve readability **Dec 19**

A minor addition was added to the bottom of the Ugly Construction web page. Flux pen photo and text. **Dec 19**

Minor updates to the VFO 2008 "Stacked Toroids" web page under **Epilogue - December 19, 2009**

Update to JavaScript applet **Item E: Calculate Cut off Frequency for an RC Low Pass Filter**. Now has a capacitor range from 0.1 nF to 1500 nF. QRP tools page. **Dec 19**

Editing plus a photo added to the broadband transformer page. This page was improved to support an upcoming project. **Dec 17**

Drafting errors on this schematic corrected (fuse position+ negative rail LED polarity). From this web page. Thanks to Paul, K0EET and Tom for the good eyes and their emails. **Dec 16**

VFO Experiments 2009 updated again; 3 images added. Supplemental web page added and updated **Dec 13**

New JavaScript applet added to QRP Tools page; **Item F: dBm calculator. Dec 12**

Nov 22-27, 2009

1 Hertz Precision Time Base added to the Hobbyist Page.
VFO Experiments 2009 added
The top level menu item "Java Tools" was renamed **QRP Tools.** This menu provides a link to my Webmaster's page and some basic JavaScript applets. Some new material will appear on the Junk Box page during this time.

Nov 20, 2009

Finally; Fall-Winter experiments begin. A small update on the VFO 2008 web page under A 3.5 MHz VFO for Diode Ring Mixers was added.

The RF Preamp page was updated with some content that was first presented on the Junk Box web page. I have received a lot of email about these amps and decided to permanently add them to the site.

The W7ZOI file linked on this page is now in pdf format.

Nov 1, 2009

The Ugly Construction web page was augmented and re-written. 2 new photos were added. This term actually came from Wes and Roger Hayward.

A new Electronic Hobbyist page was added to the drop down 2009 menu.

Oct 25, 2009

Some small updates to the Junk Box page were made. I don't ever think I have been so excited about upcoming Fall and Winter experiments as there are a number of cool, new ideas in my notebook. Extra work and travelling have kept me off the bench, but this will cease in mid November. After that, it's back to the work bench. The Fall-Winter experiments will include some HAM, SWL, general electronics and tube guitar amp experiments. Thanks for your feedback and ideas!

Oct 19, 2009

RSS feed. Click on the **orange RSS icon** above to establish a feed. I will only show 1 item; the latest addition of major new content to the site. Additions to the Junk Box page will occasionally be counted as "new content" and will be included on the feed.

Can anyone guess what brand of beer is in the tool box ? Yes.... its

Oct 12, 2009

Fall weather has come to Western North America! Thanks to Cor, PA3COR for debugging the JavaScript code on this page. Apps number D and E now work in Firefox. Rediscover the fun and learning of scratch homebrew electronics!

RF — Test and Measurement

About My Web Site

Welcome Friends!

Introduction

Welcome to the QRP and SWL Homebuilder web site. I write about my experiments with relatively simple and primitive electronic circuits.

Avoiding excessive algebra and obscure parts, I emphasize and show fundamental bench practices.

Through real experiments I examine topics to challenge and intrigue amateur designers — providing examples and describing ways to plan, problem solve, breadboard and measure your circuits.

As amateur experimenters we ought to advance in our hobby; not just perform cookbook electronics. Designing and improving your circuits requires considerable knowledge and effort. Fortunately, others selflessly share their ideas to teach us.

In time, you may recognize your electronics workbench as your greatest teacher. Bench experiments involve us thinking about and measuring our circuits so we know what's happening instead of relying too much on folklore, guessing and copying others. Designing and/or simulating circuits with software can enhance your learning but does not obviate the need to spend time in the trenches with meters, wires and solder.

People often learn skills by modeling others. We need sound examples of how other builders work and think to inform our own designs — inspired, creative and active learning driven by experience and reflection.

At some level, our bench experiences are stories of growth and realization sparked by going and doing. For example, why did the designer choose a particular resistor value? You try different resistor values while measuring the results and increase your knowledge. Collecting schematics, kit building or just thinking do not provide as intimate a learning experience as soldering your ideas on a bare copper board. Talking, tweeting or day dreaming about design is not the same as *doing it*.

Russian novelist Fyodor Dostoyevsky describes the contrast between real life and passivism; "love in action is a harsh and dreadful thing compared to love in dreams", The Brothers Karamazov, Братья Карамазовы. Dare to dream, but better yet, dare to innovate — to design and build *your own* circuits. You may start by just modifying a favorite circuit or scaling a stage to another frequency. We need more innovators and less imitators to grow and sustain our great hobby.

Electronic design produces more than a completed circuit. On the bench, even joy is experiential – a moment of discovery (or several discrete moments) yields more pleasure than stuffing a circuit board or operating a piece of gear. Creativity trumps process every time!

I hope this site demonstrates my passion for building basic, "popcorn" circuits and sharing ideas. Please remember I am just a lay person experimenter and not an electrical engineer.

Regards, Todd, VE7BPO

Essay for 2010

Building or buying test equipment and acquiring a good reference library are important to your experiments. Spurred by the realization that sound bench measurement practices are at the heart of good design, test equipment receives greater focus in 2010 and on.

A reference library is vital to our electronic experiments; good examples lead to better experiments. Poor circuits are everywhere and some builders can't tell a good design from a bad one. Minimalism and simplicity aren't excuses for sloppy design when your goal is to learn. Collecting and sharing well designed circuits helps us avoid wasting time and experiencing frustration. Circuits with attributes like well defined input or output impedances, low noise or harmonic distortion are desirable to fuel experimentation. Look for better quality circuit examples in 2010 and on.

The Internet is changing how we read and write. The prevalence of small portable web devices such as iPhones, ever increasing numbers of web sites and blogs, and the use of search engines create fierce competition among sites. Modern sites attract your attention with varied visual, aural and textural media and unfortunately, hype and pseudo-journalism. Narrative writing is more skimmed than read. Brief is in — bullets, subtitles, lists and graphics replace long lines of narrative prose which no one seems to have time for anymore.

Have you noticed the changes on this website? New content still contains lots of narrative writing, but assumes an active voice, with emphasis on brevity, clarity and speaking directly to you, the reader. Sharing mostly obsolete, analog 1970's-style circuits, QRP/SWL HomeBuilder attracts a tiny, niche audience. I believe the success of this website depends on providing good and diverse content — not Tags, RS feeds, adopting net-speak, or self promotion. You be the judge.

Essay for 2014

The Internet of Everything?

Bucking the trend, my contribution to amateur RF homebrew remains informational and not social. Why?

Social media information represents a Pandora's Box of good science and opinion, mediocre thought, or trash potentially created and/or disseminated by anyone who's connected. We accept that much of our social media content doesn't come from the best or brightest — some people are just plain interesting, or express themselves vigorously, seem like-minded, or touch our hearts.

Some builders, like me, seek objectivity and not just "likes" and "follows" based on sentiment and spectacle. While a few radio builders may prefer to join hands and sing Kumbaya, or pat themselves and others on the back simply because their breadboard actually works, a trifling of us care more about how and why our circuits work. We *like* measures and measurement tools and *follow* science, experiments and the works of those who shine brightly.

What's wrong with plain information, unfettered discovery, experimental rigor, objectivity and rational, kind thought?

It's not that these characteristics don't attribute social web clients — they do, but the negative impacts of social media worry me a little. A brief list of concerns: loss of privacy, the threat of wasting time while really just isolating ourselves from our real friends + family. The numbing exposure to the Internet of Ads and Spam. Still too, bubbling up like purulent sores come the charlatans, the misinformation peddlers, the opinion spammers, and those who anonymously leave stinging sarcasm, or outright hatred [hostile online comments that attack people, or divert a healthy flow of ideas].

Running a low-tech web site with nearly 0 commercials suits me better.

SEO — Search Engine Optimization

I've read that Google analyses your web site content, the number and quality of the sites that link to your pages, their search engine clicks and so forth. In part, Google seems to rank a site based on how relevant and authoritative they believe it is. Some people specifically employ SEO techniques to gather in more traffic.
To my surprise, each year, tens of thousands come to this site via search engines like Google. I don't think my material seems too relevant or authoritative. I've made no effort at SEO, so I conclude that you, my readers have more to do with the site's success than anything I've ever done.

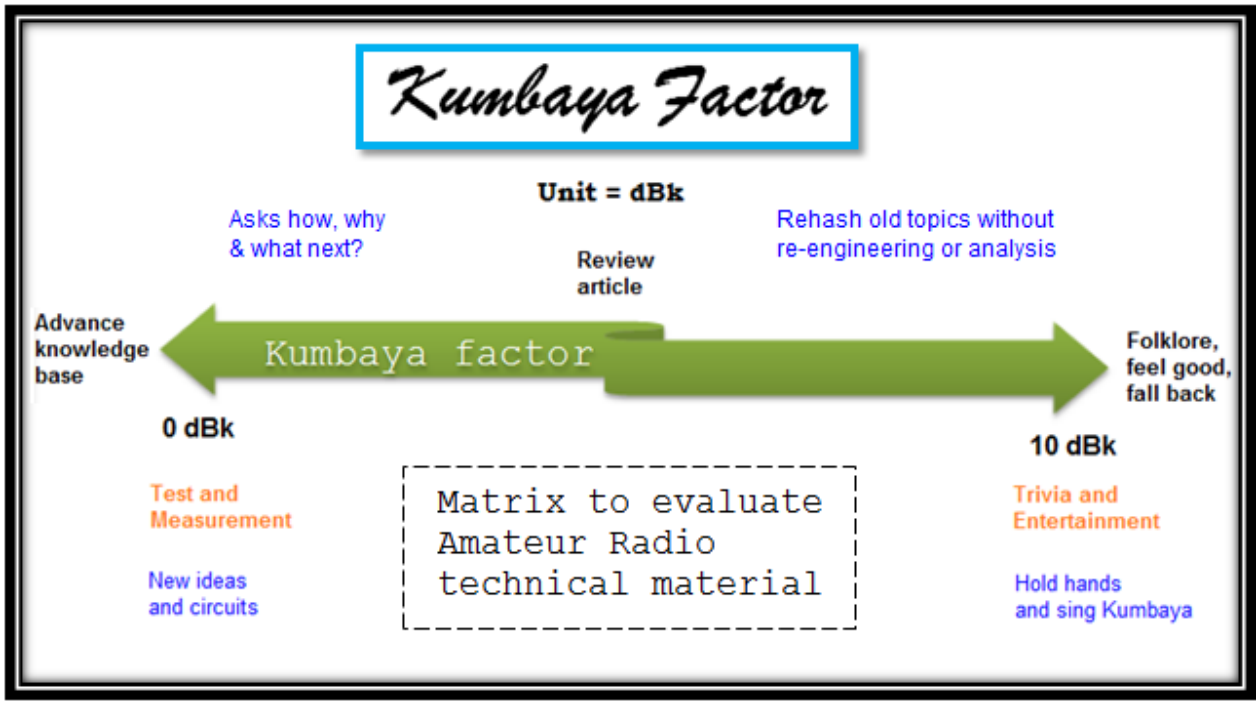
Thank you.

Hope Invigorates

Invigorated by the excellent work either emailed to me by experimenters such as Michel F6FEO, or Dick, N4HAY; or posted on blogs or Community sites like Yahoo, I feel hopeful about the future of our hobby in 2014. The PHSNA Yahoo group leaders, the recent work of Jason, NT7S, Steve VE7SL and many others show that *amateur design* experiments still have a pulse + respirations.
The aforementioned get my vote for their MOF like behavior: a strong blend of creativity, tradition and quality.

Looking Ahead — Future Site Content

Most of my new receiver work involves quadrature and in-phase mixers fed with (2) local oscillators; 1 output shifted 90 degrees from the other — essentially, EMRFD Chapter 9. Even my Funster [a personal, lowbrow trans-receiver I drag onto hill and dale] now contains phasing receiver circuitry to reduce the opposite sideband by 20 dB along with further low-pass filtering. I hope to add some Funster content to HF Embarcadero in Winter 2014. Like many of you, the 1 resource I lack the most is time.
While I'm thrilled with the notion of a receiver appliance that contains just an antenna, LNA, ADC and some sort of "wonderfall" display or speaker, I'm still smitten by analog design with hardware. Still, the I-Q mixer will offer a nice transition into SDR should I ever wish to spend my free time writing C# and not melting solder.



Best to you!

Miscellaneous

My special thanks to Wes Hayward, W7ZOI for his generous support and elmering over many years.

EMRFD is the main reference of my site

All permanant content circuits were built and tested. Schematics are drafted as carefully as possible. Please accept that bench and/or drafting errors may occur. No liability arising from the application, use, or misuse of these projects that results in direct or indirect damage or loss is assumed.

Full price is paid for all parts used and no monies are or were received for promoting any products or companies on this web site. Any ads, hyperlinks or mention of commercial products or companies is out of courtesy only.

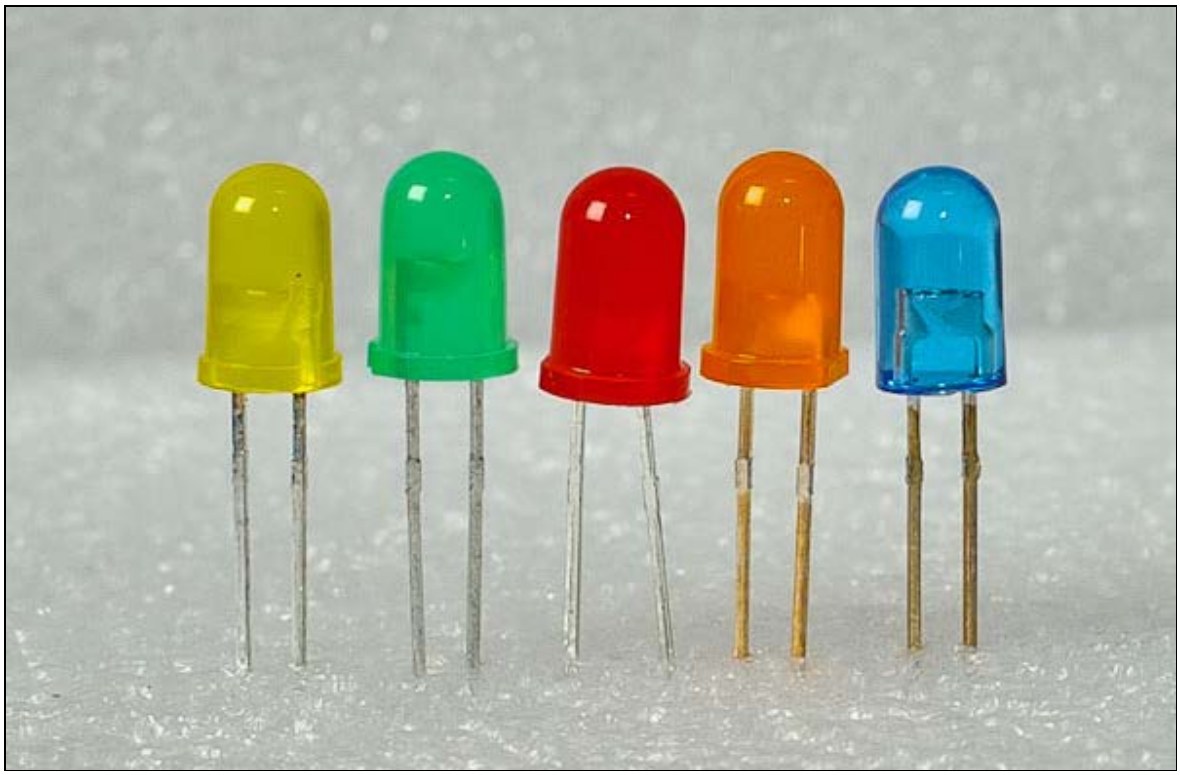
"Until you build and measure it, you don't know what you don't know";
Rick Campbell, KK7B; VHF Open Sources — Design of Low Power High-Stability Low Phase Noise Single Frequency VHF Sources with High Spectral Purity; 2008

Information Regarding the Compression of Schematics

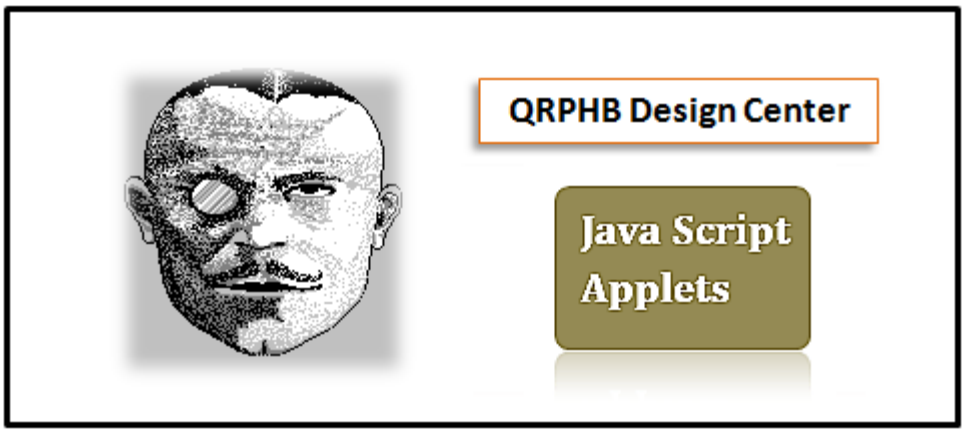
I see many electronics web authors compressing black and white schematics as jpg files. This results in distortion of the schematic. Schematics are best compressed using the 8-bit, lossless LZW algorithm which means converting the file to a png, gif, or even pdf format. The files sizes will typically be smaller than .jpg compression, have no distortion and can be edited easily.

RAC is the National Amateur Radio Society of Canada

For my web page concerning support of the **Radio Amateurs of Canada**, please click here



The hand drawn image bitmaps on this web site (logo etc.) are by Rod Adams. All website photographs were taken by VE7BPO except as indicated.



1. Click for the old QRP HomeBuilder Graphics page. Click for my Pin Outs page. Click for my Homepage

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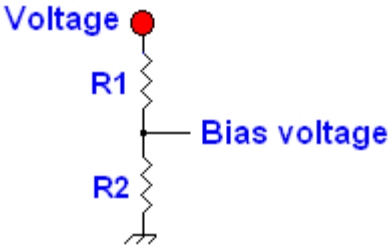
2. Some simple tools written in JavaScript for the QRP/SWL HomeBuilder:
`Minimal input error checking`

A. Calculate DC Voltage Divider Bias

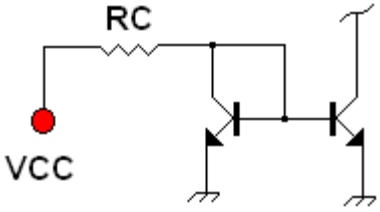


Enter Voltage: Enter R1: Enter R2: Bias voltage =

B. Calculate Inductive Reactance

Enter Inductance in uH: Enter Freq in MHz: XL (ohms) =

C. Calculate DC Current for a Current Mirror



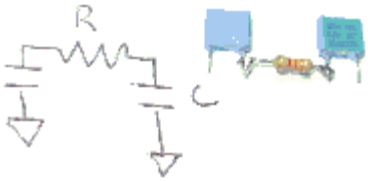
Enter VCC: Enter RC in ohms Current (mA) =

D. Calculate # of Turns To Obtain a Desired Inductance on a Ferrite Torroid

Enter Inductance in millihenries: Select Core: FT37-43 Turns =

The *AL for this ferrite core is* =

E. Calculate Cut off Frequency for an RC Low Pass Filter



Enter resistor value in ohms: Select capacitor values in uF: 0.01 3 dB down frequency (Hertz) =

F. Calculate Power in dBm and mW from Peak to Peak Voltage

Enter measured peak-to-peak voltage into a 50 ohm load: dBm, mW

Application Note: This web site follows the EMRFD standard for dBm power measurement. dBm = the power delivered into a 50 Ohm resistive load which is temporarily substituted at that point in the signal chain.

G. Calculate Lm and Cm For a Crystal using the G3UUR Method

Enter frequency in **MHz** written on crystal (series resonant frequency):

Enter measured frequency in **MHz** with switch open: Enter measured frequency in **MHz** with switch thrown:

Enter crystal capacitance in **pF**: Enter open switch circuit capacitance in **pF**:

Cm = femto Farads , **Lm** = Henries

Supplemental web page for this applet: Crystal filter measurement and adjustment Link

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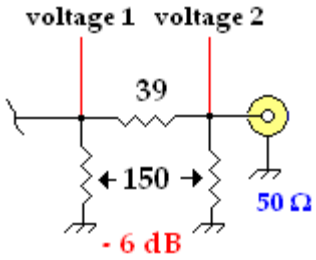
H. Calculate Decibel Power Gain or Loss from 2 Peak-peak Voltages



Enter voltage 1: Enter voltage 2 dB gain or loss =

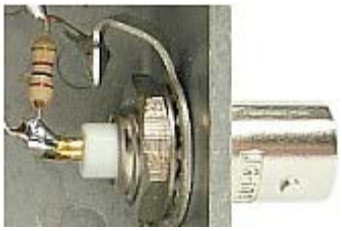
I. Calculate Decibel Power Gain or Loss from Input and Output power

Enter input power in watts: Enter output power in watts dB gain or loss =

J. Calculate dBm and mW from RMS Voltage (50 ohms)

Enter RMS voltage dBm mW

K. Calculate Return Loss and VSWR (50 Ohms Detector)



Enter the detector signal in pk-pk volts when the unknown port is terminated in an open circuit: Enter the detector signal in pk-pk volts when the unknown port is terminated in the unknown impedance:

Return loss = dB, VSWR=1:

Measurements per Figure 7.41 EMRFD. Schematic here

L. Calculate Power from the DC Output of an AD8307 Meter

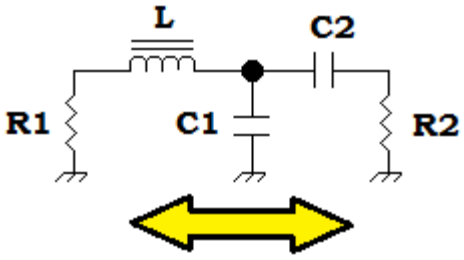
1. Linear calibration steps:

Enter measured DVM voltage at -10 dBm: Enter measured DVM voltage at -20 dBm:

2. Calculate power in dBm from DVM voltage:

Enter measured DVM voltage: Power =

M. L-C-C Tee Network



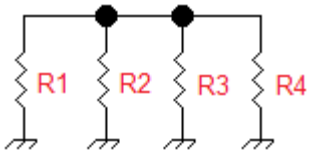
Enter **frequency** in MHz:

Enter **R1** in Ω: Enter **R2** in Ω: *R1 must be < R2, but the network is bi-directional*

Enter Q: *Perhaps start with 2-5*

C1 = pF, C2 = pF, L = uH

N. Parallel Resistor Values (2-4 resistors)



Enter R1: Enter R2 Enter R3: Enter R4:

R = ohms

O. LCR -- a Reactance Calculator In beta -- do not use !

Mode: L C to Reactance Inductance: uH Enter frequency in MHz

This page last updated: September 21, 2013

RF — Test and Measurement

Short Wave Listening коротковолновое радио

Introduction to Short and Medium Wave Radio Listening

Short wave radio listening was a childhood passion and I enjoy being an SWL just as much today and log at least 800 hours of SWL per year. There seems to be many web pages devoted to construction of radio equipment for the amateur radio experimenter but relatively few for the shortwave radio devotee. I decided to expand this web site to include projects for the SWL Homebuilder in 2005.

My favorite bands are 49 meters (5.9 - 6.2 MHz) at night-time and 19 meters (15.1-15.8 MHz) during the daylight hours. I also listen to medium wave DX around 1400 - 1600 KHz.

Why Listen to Analog Short Wave Radio?

Is analog short wave radio dead? I think not.

No doubt, short wave radio has passed its prime and is slowly dying, however, it's still fun and/or relevant to some.

World band radio: Almost 1/2 of the world's population lives on $2.00 *USD* or less per day. The Internet (the main alternative to shortwave radio) poses a luxury to many poor people living in lower-income countries — experienced travelers or those who support people in developing countries will understand this statement. In some countries now, ranking in the middle class just means you have a full-time job. In addition, oppressive governmental regimes may limit foreign media and Internet access: LW, MW and SW radio can break through obstacles such as natural or man-made disasters, borders, poverty and censorship.

For SWL hobbyists, analog shortwave radio entertains, informs and best of all, provides opportunities to analyze propagation and experiment with real radio topics including static, solar flares, QRN, antennas, grounding, baluns, coax, and wire. SWLing poses an adventure — it's unpredictable, challenging and increasingly difficult as stations decrease and QRN increases. I've built many antennas and even some noise cancelling circuits just to pull in a few Dx stations. The sport of SWL lies in making DX contacts: a theme shared with Ham radio.

What About Internet Radio?

Radio by definition is the transmission and reception of electromagnetic waves of **radio frequency**; but perhaps blue-tooth or Wi-Fi reception from a hot spot qualifies as *radio* in the modern era? Just as peanuts aren't nuts, Internet server or webcasted radio is not RF broadcasted radio. I think Internet radio is great, but fundamentally a very different medium from that enjoyed by SWL fans

Internet radio involves a radio player decoding a stream of compressed bits fed from a Internet radio station or virtual receiver. In some cases, the material originates from a real radio station that also broadcasts an AM or FM signal. For example, you can tune FM station Rooskie Radio "Русское Радио" in much of Slavik Europe or play them on a computer device anywhere you can get an Internet connection.

For lovers of foreign content, listening to Internet radio makes sense; providing convenience, a good signal when bandwidth is high and 24 hour per day listening on 1 IP address. Internet radio offers a much cheaper way for content providers to beam their news and music services around the globe — we've seen numerous large broadcast radio services such as the BBC World Service reduce or drop analog SW and add Internet radio, satellite and digital SW transmissions for their customers.

The exciting growth of independent and niche Internet radio stations increases personal freedom of choice and provides opportunities for unique

interest providers and consumers to find each. Media streaming companies and manufacturers of Internet radio *players* and their worldwide distributors benefit too.

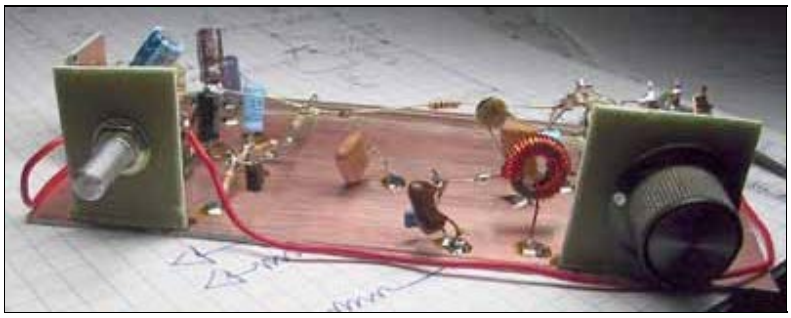
This technology is a far cry from tuning the SW bands with a homebrew or commercial radio frequency *receiver* and a length of wire slung in a tree. Perhaps, the greatest advantages of Internet radio are that you don't have to get up early, or stay up late to pull in some rare Dx, nor do you need any radio skills or special equipment — perfect for the majority of listeners. But we're SWL radio hobbyists: people who listen for both content and because we love radio propagation and gear.

There is nothing wrong with Internet radio, or any of the modern data streaming techniques however, SWL aficionados driven by skill, the thrill of Dx and love of their experimental hobby share a special bond that Internet radio doesn't give them.

Assembling a station The most important component in your radio shack is your antenna. Don't hesitate to safely experiment with the many antenna designs available on the world wide web. Your sure to find a commercial unit or home brew antenna design that suits your real estate and budget.

Your next task is to find a receiver. It is difficult to recommend any one receiver because there are so many excellent commercial receivers to choose from. If you are thinking about purchasing a used receiver, you might consider checking eBay to find a receiver or to learn the going price for used gear. The ultimate SWL experience in my opinion is to build and operate a receiver on at least 1 band.

Favorite SWL and SWL-related Web Sites



Wikipedia-Shortwave Bands A good description of the bands and their general propagation.

Canada's SWL-DXer website Hard core Canadian web site dedicated to SWL. Thanks gentlemen!

http://www.bobsamerica.com/swl

http://www.dxing.info/

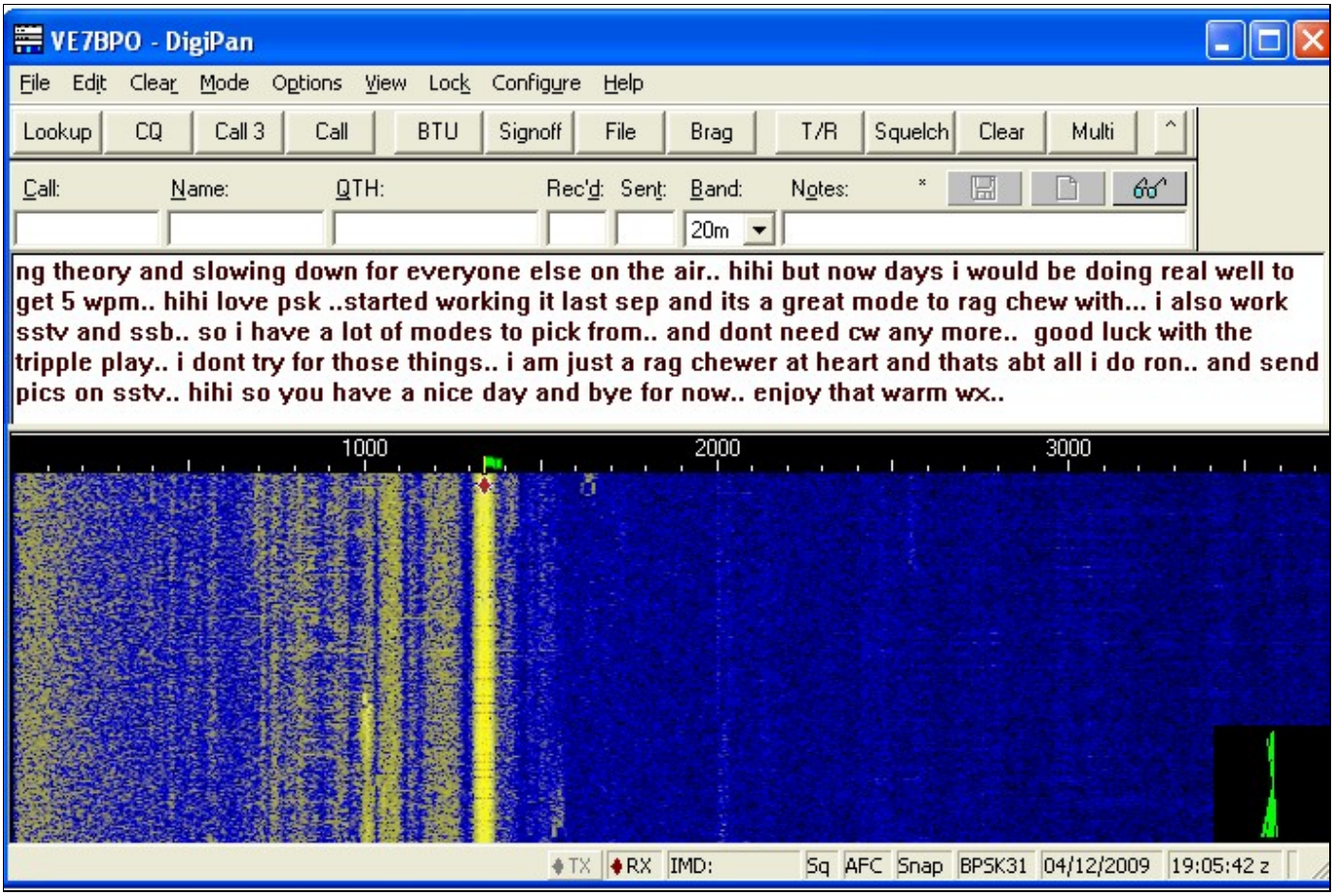
Doug's Shortwave Radio Page

AA6V's SWL Links

Method for soldering a PL-259 to RG-213 or RG-8

Digital Modes For SWL Fans

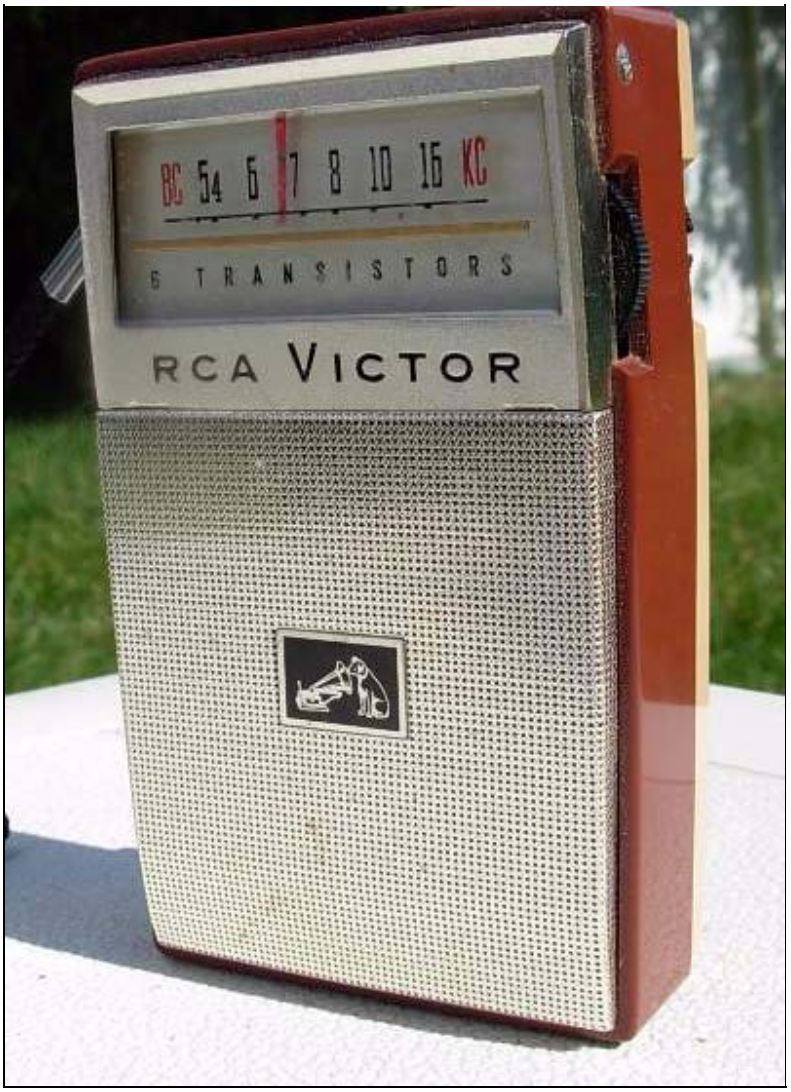
PSK31



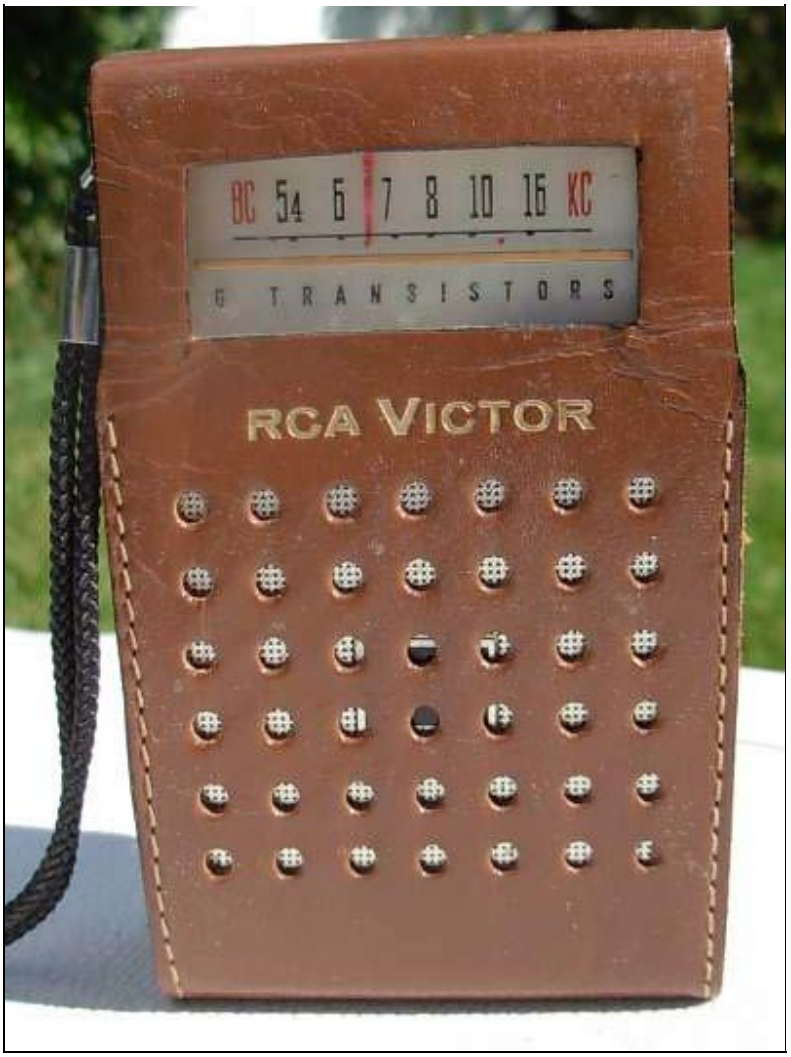
There are a number of good sites about this relatively new HAM mode. All that you minimally need is a receiver dialed in at 14.070 USB (or another

PSK31 frequency), a microphone hooked to your computer sound card and some free software. The software (DigiPan 2.0) is available at http://www.digipan.net/ .
I use a USB interfaced microphone and place it about 2 cm from my receiver speaker. If HAMs are operating; you should hear some warbles and see some waterfalls on your screen. Click on one of the waterfalls to begin receiving the text. I knew nothing about this mode, but was up and running in 15 minutes.

Additional Short Wave and Medium Wave Receiver Photos


BC 54 6 7 8 10 15 KC
6 TRANSISTORS
RCA VICTOR

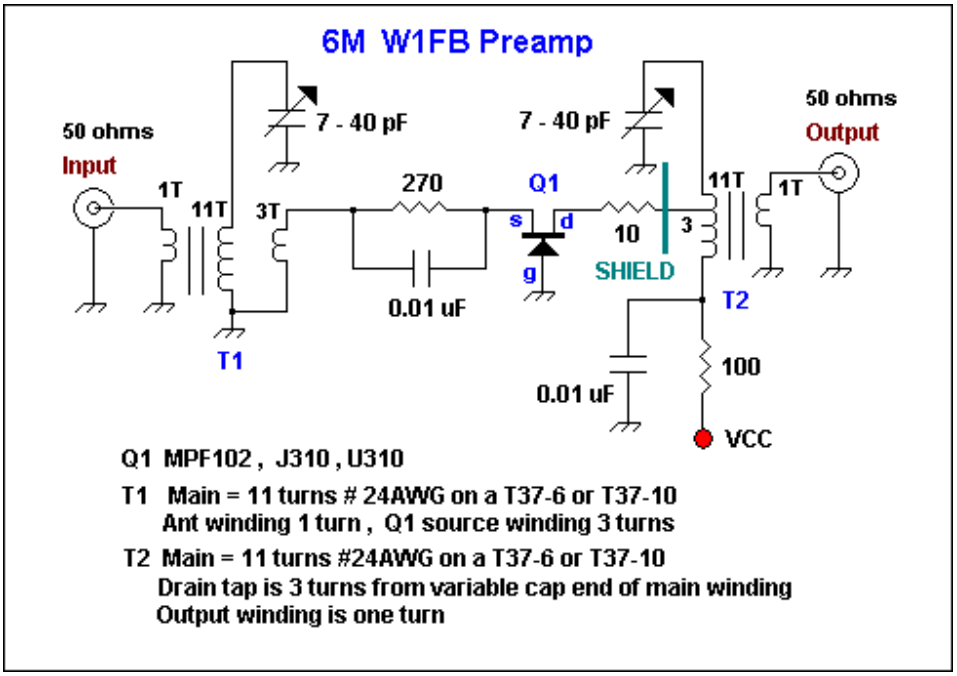
RCA VICTOR
SISTOR
DIO
MODEL 6P-626
RCA VICTOR
RCA VICTOR


BC 54 6 7 8 10 16 KC
6 TRANSISTORS
RCA VICTOR

RF Preamps

W1FB 6M RF Preamp

Discussion:

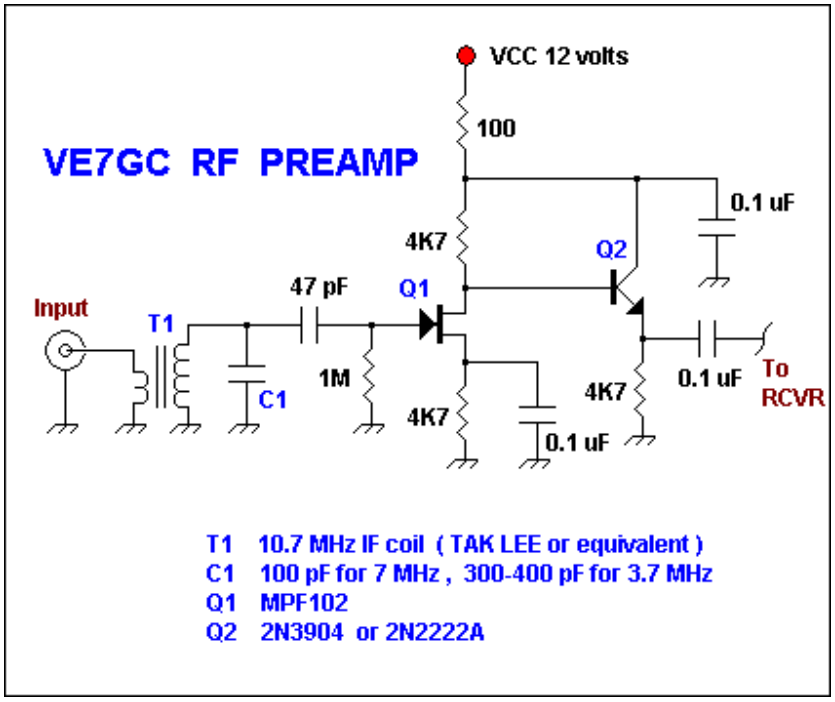


Here is a schematic sent to me by W1FB many years ago. It is very similar to a 6M two-stage preamp that he published in QST in the mid eighties. Doug really favored the grounded gate FET for narrow band preamps. His published work is replete with examples of them on just about every band. I built that amp and remember getting about 10 dB gain, which is all that I wanted for the 6M direct conversion receiver using a diode ring detector that I was building. The great feature of the amp is that it combines a band pass filter and preamp in one. I lost the original schematic that Doug sent me but was delighted to see that I made a bitmapped drawing of it on a floppy disk that was recently re-discovered when we were moving an old desk. The shield shown in the schematic was a small piece of grounded ,double sided PC board in which, I made a small chamfered hole in to pass the lead going to the T2 tap. The shield, along with very short component leads will help minimize parasitic oscillations. The T2 tap is 3 turns down from the end of the T2 main winding that connects to the variable capacitor. Doug specified T37-10 cores for the inductors, but I substituted T37-6 cores and used the same number of windings as specified for the T37-10 core inductors. It worked fine.

VE7GC Popcorn RF Preamp

Discussion:

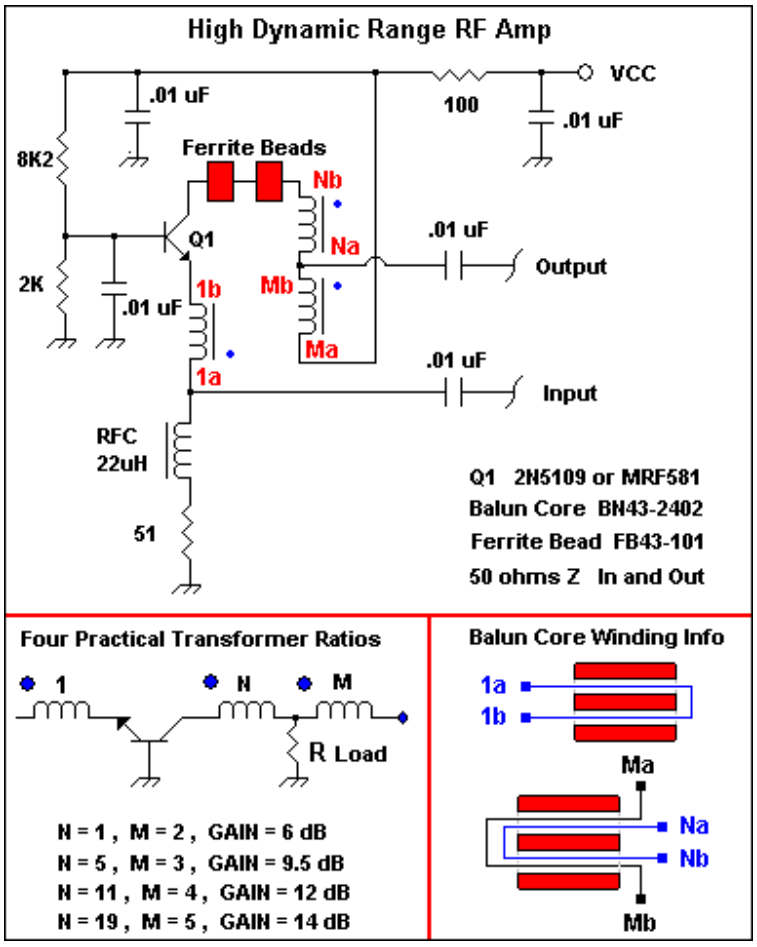
Here is an easy RF preamp by Dick Pattinson, VE7GC. It uses a single tuned circuit at the front end and can connect directly to a mixer or product detector in a simple receiver project. Note how Dick provided adjustable RF gain control for this circuit in his Wee Willy project on this website. If you can not find Tak Lee green 10.7 MHz IF coils, probably any other brand of 10.7 MHz slug tuned IF transformer would work. The Mouser catalog number is 421F123 . If your 10.7 MHz IF coil has a built in capacitor at the base , remove it. A fixed inductor may also be wound using a powdered iron torroid core and then all or a portion of the C1 capacity would be made variable. The input impedance is 50 ohms and the output impedance is low due to the Q2 emitter follower stage.



A Low Noise, High Dynamic Range Broadband RF Amp

Discussion:

This schematic is a version of a circuit developed and patented by David Norton and Allen Podell in June 1974. This variation was described by Joe Reisert, W1JR in the now defunct Ham Radio Magazine. The Norton design uses transformer coupling to achieve "noiseless negative feedback" and is really outstanding.
A great article utilizing and augmenting on this technique receivers is by Jacob Makhinson, N6NWP in QST magazine for Feb 1993 with "A High Dynamic Range MF/HF Receiver Front End". Makhinson arranged 2 in push-pull to obtain excellent results. Obtain a back-issue of QST for closer study. **Note** that the fore mentioned Feb QST article has the coil phasing wrong and the correct phasing can be seen at this web site from QST for July 1996. There is also information about Norton feedback RF amplifiers in EMFRD.



If you are building a contest-grade receiver and need a good RF preamp and/or post mixer amplifier, the Norton type is quite suitable. An amp built using a 2N5109 can have a noise figure in the 2.5 - 3dB range. I have also built them with 2N3866, MRF517, MRF581 and a 2N5179 although the last transistor would be a somewhat poorer choice. This schematic with a 2N5109 is good from 1.8 to 150 MHz with a 1.2:1 VSWR or less according to Joe Reisert. I have even put one in a friends CB radio and he was delighted.

Winding and Construction Hints

Making the Norton amps requires some planning to keep all component leads as short as possible. The transistor leads and any connecting components should be trimmed as short as practical to promote stability. Sketch the component layout on a piece of paper and modify it until you are satisfied you have designed a good layout. The ferrite beads on the transistor collector aid in stability and should be used to preserve the noise figure by squashing any oscillations should they develop. The 22 uH choke can be the little epoxy coated units that are color coded and look somewhat like resistors. Do not use a choke less than 22 uH.

Before winding, the builder must first decide how much gain is needed from the amp. For an RF preamp, the stage should have gain equal to or greater than the passive stages after it. Also there will be losses in the transformer, so the theoretical gain of the Norton amp maybe 1 dB off and will need to be factored in. For the purposes of discussion, a 9.5 dB amp is desired , so N = 5 and M = 3. The first step is to mark one side of the core with a dab of liquid paper, paint or a small piece of tape. This will allow you to keep track of the transformer later. To mark, hold the core so

that both channels are parallel to the floor, one on top of the other. Apply your dab of paint to the top of the core and use the marked top to denote the A windings. 1a, Ma and Na will all start from the top channel in the balun core. Using 32 AWG wire for all three windings, start with winding 1 and wind the single turn from point 1a to 1b. Cut off the leads so they are shorter than 5 centimeters (2 inches). Next, wind Ma to Mb three complete turns through the binocular core and trim the leads if needed. Tie a small knot in the wire at both ends. This will clearly mark this M winding. Both windings should look like the diagram under the schematic. 1a to 1b are on the left of the balun core and winding Ma to Mb are on the right side of the core. Mb has a distinguishing knot at the tip of both wire ends. Ma starts from the top of the core which you have marked with a dab of paint or something. Finally, wind Na to Nb five complete turns through the core in the same direction as the previous winding M. Strip wires Na and Mb (Mb has the knot), twist together and solder. Scrape the enamel off the leads **very** gently with a sharp hobbyist knife.

Insert the transformer in your circuit and cut the leads to their proper length and then solder away. It maybe preferable to pre-strip the leads on winding 1 as it is hard to strip the enamel off a fine wire that has only one turn and it may accidentally pull out of the core. If it does, just re-insert it into the balun core on the correct side. Once you have soldered Na and Mb you can always identify the windings later because you have marked the top of the balun core which denotes the A windings. Try and make your windings gently tight as if there is too much slack you may have difficulty getting the last few windings thru the core channels. A 14 dB gain amp maybe **impossible** to wind with 32 AWG wire, it may best to use 34 AWG for that amplifier. I have never built one for greater than 12 dB. The transformers are a bit tedious to wind, however persevere and the results will be well worth it. For HF, you can substitute 0.1 uF caps for the 0.01 caps shown if you like.

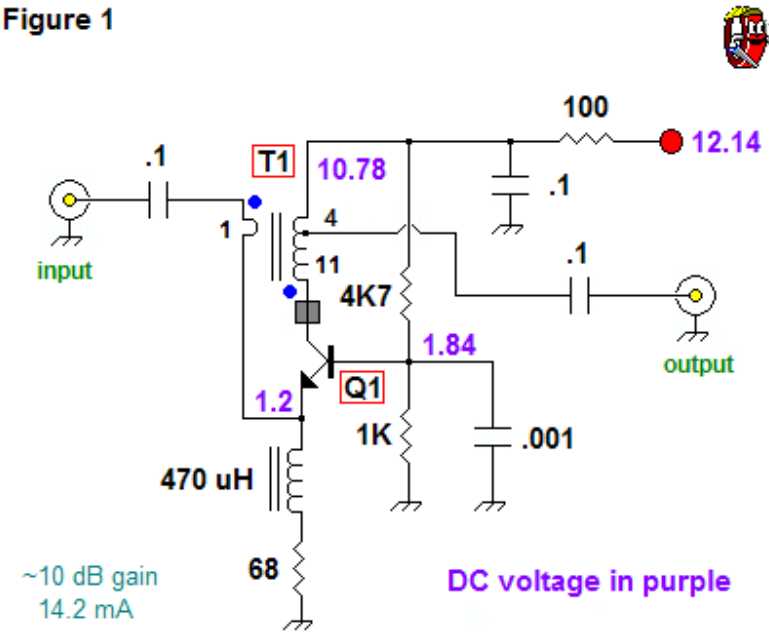
Toroidal Inductor Norton Amp Experiments

Discussion:

The amp shown in the schematic to the right uses a ferrite torroid for the transformer and has ~10 dB gain. Winding 1 turn of wire over the cold end as shown in the schematic is tricky. Try to keep this link as short as possible. A ferrite bead or a 22-51 ohm resistor on the transistor collector is desirable. You can try increasing the turns (1:21:5 etc) to experimentally obtain more gain from this amp. The torroid version is a valid option for builders who do not have balun (binocular) core ferrites in their junk box. Toroidal inductors are certainly easier to wind then binocular core versions.

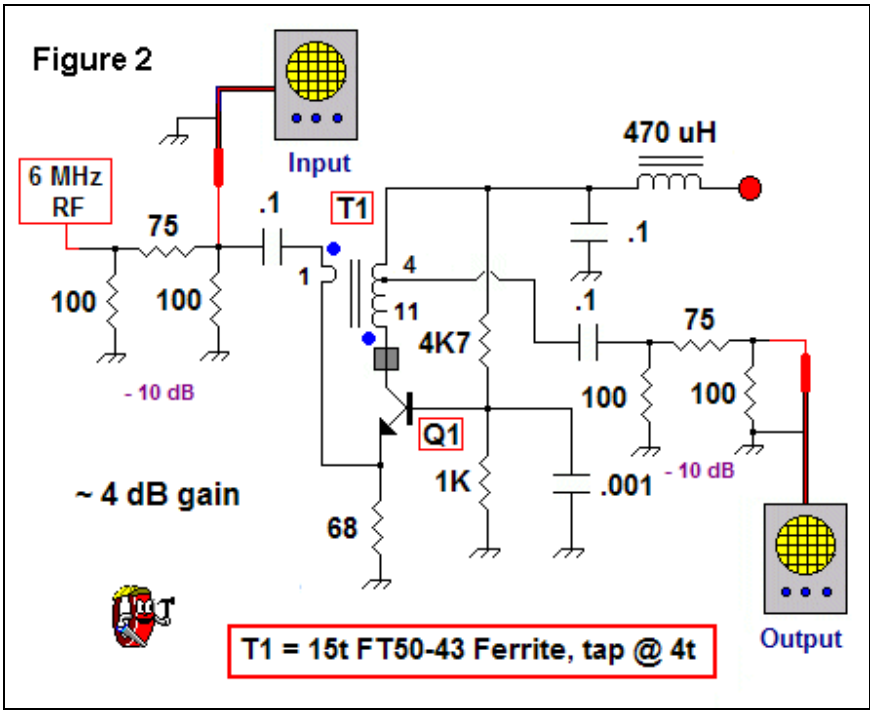
In 2007, I built several Norton "noiseless feedback" RF amps using FT50-43 and FT37-43 ferrite torroids. These are outstanding and I recommend using them in projects. The input and output Z is 50 ohms. The overall BJT topology is reminiscent of a common base amp. I have some basic information concerning this amp on this web page . They are straight forward to build. The biggest problem is the phasing of the single turn link. Get it wrong and your amp can turn into an oscillator.

Figure 1

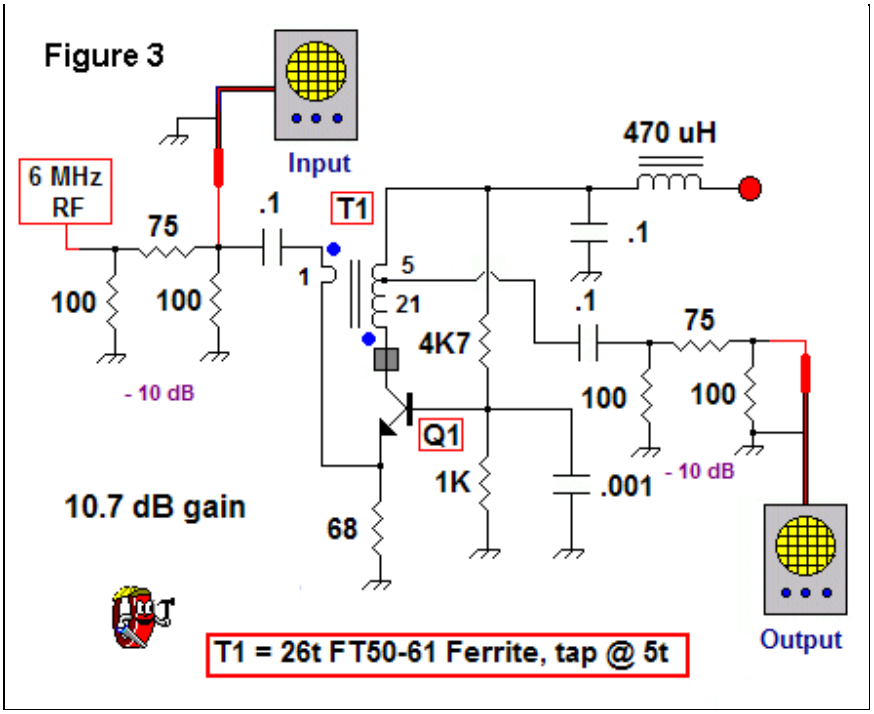


Q1 = 2N5109 or NTE 123 etc
T1 = Secondary: 15 turns with a tap a 4 turns from the cold end.
Primary: 1 link over cold end
T1: use FT50-43 or FT37-43 ferrite

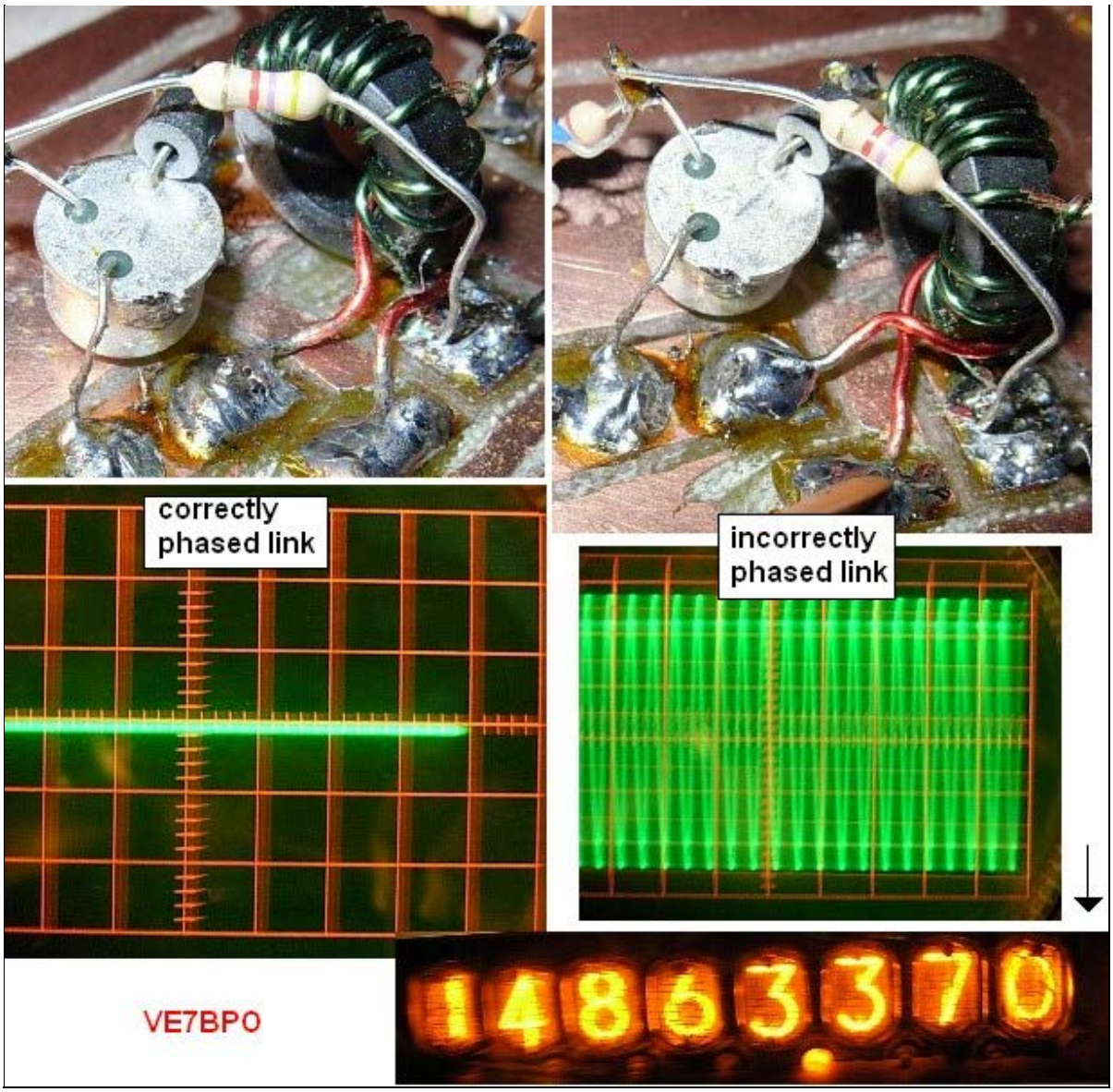
Shown above. The breadboard of Figure 1.



Shown above is the Figure 1 amp above (labeled Figure 2) with a 50 ohm -10 dB pad on the input and output, so gain is low. I used these pads to evaluate the amp in a number of experiments. I never got around to writing up these experiments on the web site and likely never will. I wish I had more time as my notebooks are full of unpublished experiments that would be great content for this web site.

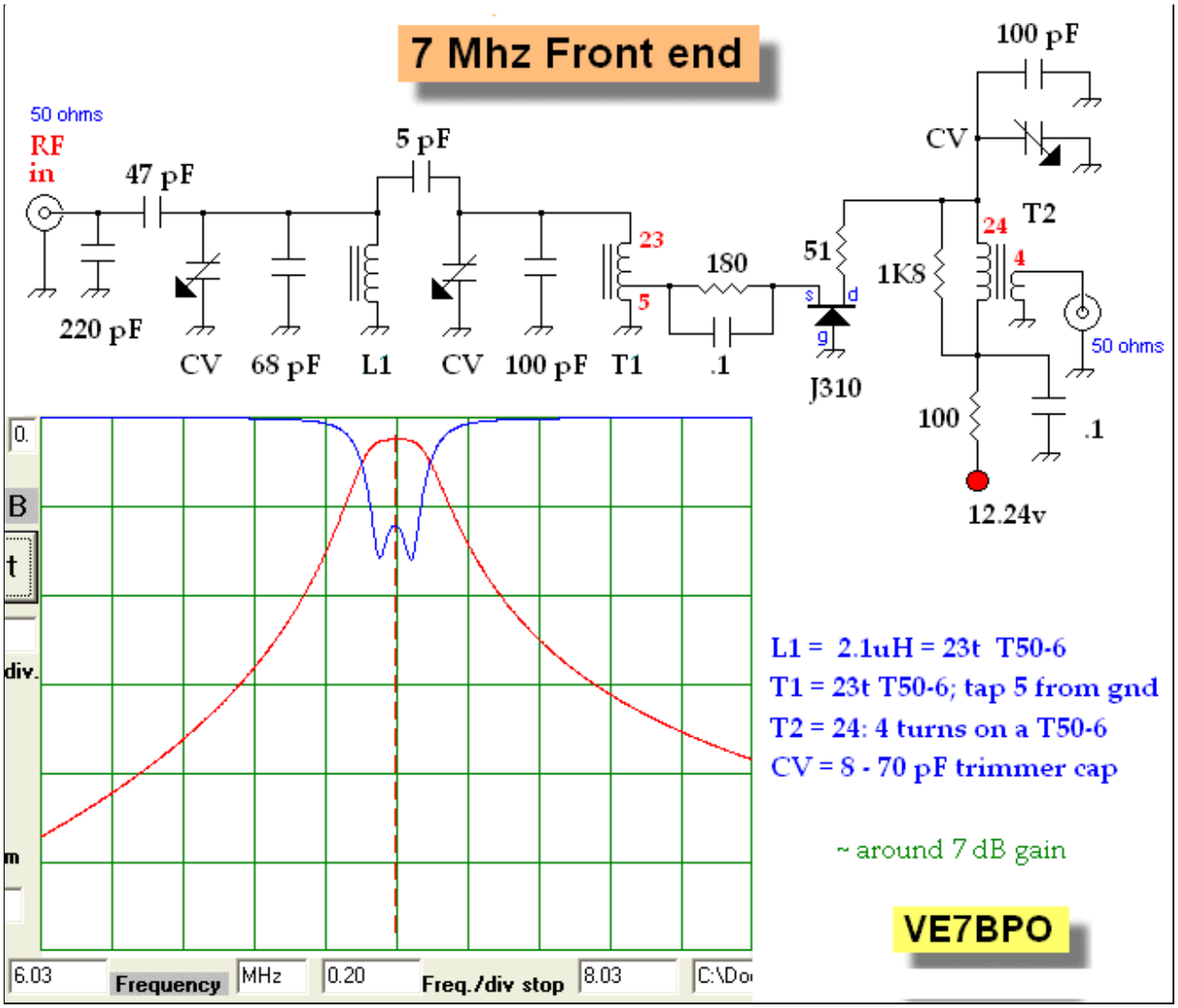


The amp above (labeled Figure 3) is a hot one; 10.7 dB gain even with 10 dB of attenuation. You can leave off the input pad and decrease the output pad to -6dB if you want or require a wide band, low noise RF amp with lots of guts. Most builders use binocular ferrite cores for the inductor, but torroids work fine for many applications.

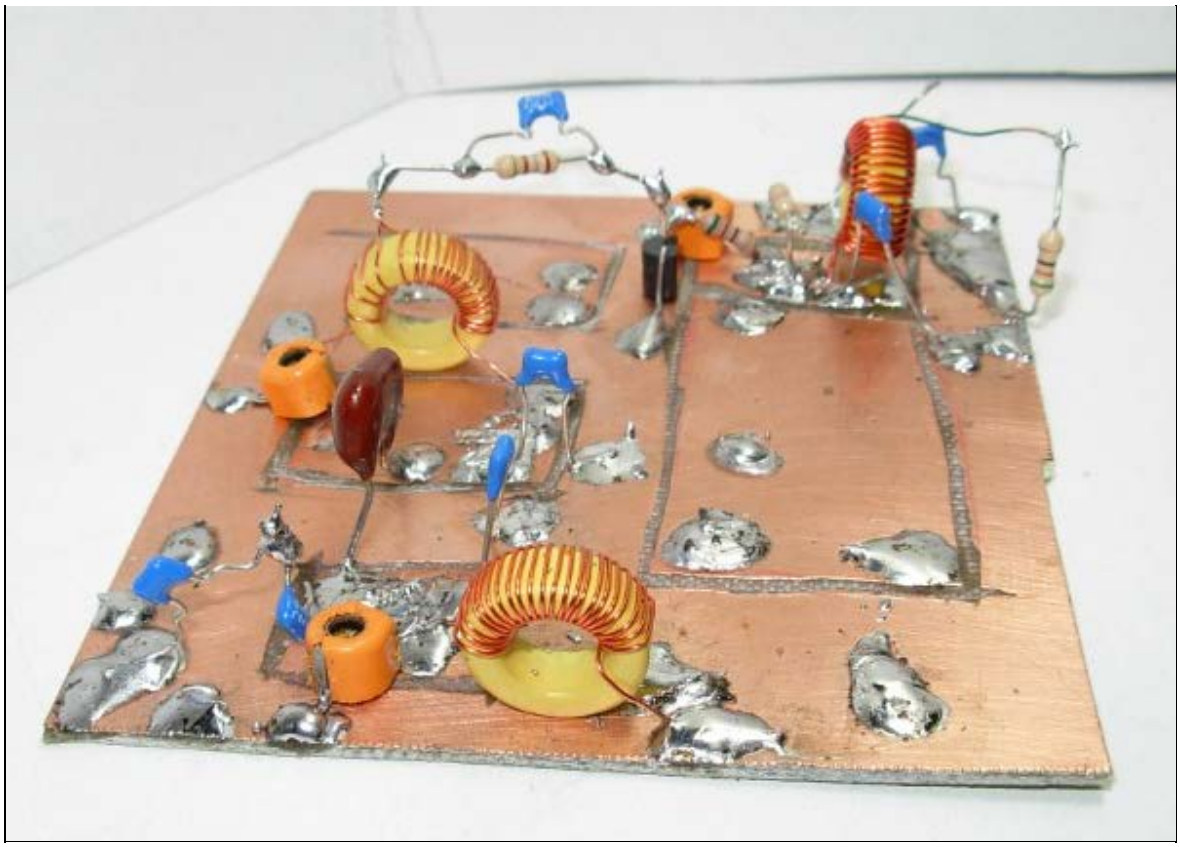


Shown above is a photograph of 1 of the experiments from 2007. The one turn link from the Norton amplifiers just above is shown phased correctly and then phased incorrectly. Note the oscillation in the "badly wired" amp at 14.86 MHz. I routinely check all of my noiseless RF amps using the oscilloscope. Occasionally, I will put a shunt coil and cap (from input to ground) on the input to "exaggerate" any oscillations. This has proven to be a useful technique for testing if the phasing of the one turn link was done correctly.

RF preamp for the 40 Meter band with 3 tuned filters



An experimenter's 40 Meter band front end for CW. This has a double-tuned filter and a low gain, lower noise RF amp. Great circuit for isolation of a product detector or mixer in a popcorn receiver.



A photograph of the above 40 meter band front end, double-tuned filter plus tuned common gate RF amplifier. Input and output Z is 50 ohms.

Amateur and Short Wave Radio Electronics Experimenter's Web Site

QRP HomeBuilder Software

Disclaimer:

THE QRP HOMEBUILDER SOFTWARE OFFERED HEREIN ("THE SOFTWARE") DOES NOT COME WITH ANY WARRANTY, EXPRESS OR IMPLIED. IF YOU MAKE USE OF THE SOFTWARE, PLEASE BE AWARE THAT YOU DO SO AT YOUR OWN RISK. NEITHER THE AUTHORS OF THE SOFTWARE AT JENNA DESIGN NOR ANY OTHER PARTY WILL ACCEPT RESPONSIBILITY FOR ANY OCURRING OR UNFORESEEN CONSEQUENCES OR DAMAGES THAT ARISE AS A RESULT OF THE USE OR MISUSE OF THE SOFTWARE.

Technical Info and Distribution

The QRP HomeBuilder applications are written in C++ for speed and compactness. Apps will specified as GUI (graphical user interface) or 32-bit console based (DOS look). Anyone may display or distribute these applications via website or diskette providing that they do not charge for the program(s). I know longer have a C++ compiler and no future work on these applications is anticipated.

QRP - RELATED APPLICATIONS FOR DOWNLOADING

CoilBuilder_99

CoilBuilder_99 is a powdered iron inductor winding application. Enter desired inductance, select core size and mix and press the Calculate button to determine the correct number of windings for your inductor. Data is also given showing, core color, permeability, frequency range, AL value and maximum number of turns versus wire guage for the chosen core size. Encompasses 12 different core sizes and 8 different mixes of powdered iron. Calculated results can be stored on a disk file or printed out.

Style: GUI, File size: 90K, zipped, 44K.

Bug Fixes: Some missing AL values for # 7 material added April 24/99. K6WHP's superior version is linked below.

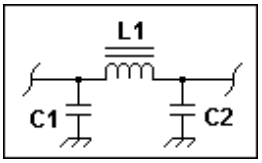
Current Version is: 4 / 24 / 1999

Download the CB99.zip file

PI Filter Designer

PI Filter Designer is a simple 3 element 50 ohm input and output impedance pi filter designing application. This program allows the user to design simple lowpass filters by selecting from a variety of standard capacitor values either empirically or to suit what you have on hand. The filter 3 dB cutoff frequency and required L1 inductance are automatically calculated and displayed. In addition, the user may select an additional capacitor value to put in parallel with both caps C1 and C2. In this app XL = XC = 50 ohms impedance. No other impedances can be calculated with this program.

Style: GUI, File size: 47K, zipped, 22K.



Current Version is: 1 / 14 / 1999

Download the pifilter.zip file

CapCoder

CapCoder gives the capacitance in microfarads, nanofarads and picofarads and tolerance of any capacitor code entered into its input section. Example : 104J. This app uses numeric spin-buttons and a combo box so that no typing is required for data entry.

Style: GUI, File size: 48K, zipped, 22K.

Bug Fixes : A nanofarad conversion error was corrected July 2, 1999.

Current Version is: 07 / 02 / 1999

Download the capcoder.zip file

Resistor Coder

Resistor Coder gives the resistance in ohms of any resistor color code entered into its input section. Four or five band resistors can be accommodated by this program. This app uses drop-down combo boxes so that no typing is required for data entry. Results may be saved to a disk file or directly printed.

Style: GUI, File size: 58K, zipped, 27K.

Current Version is: 1 / 16 / 1999

Download the resistor coder.zip file

Ferrite

Ferrite is used to calculate the number of turns required on toroidal ferrite cores to achieve the desired millihenry-value inductance. 15 different ferrite toroids are included in this application. This program will calculate the winding data for an inductance range of 0.001 to 27 millihenries.

Style: Console, File size: 64K, zipped, 31K.

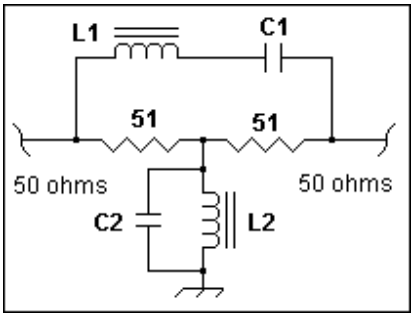
Bug Fixes: Thanks to PA3CKR for the bug report; fixed Jan 19/99.

Current Version is: 1 / 19 / 1999

Download the ferrite.zip file

Universal Diplexer

Universal Diplexer calculates the inductance and capacitance values for a Bridge-Tee diplexer based upon a chosen superhet receiver intermediate frequency. The diplexer is the Joe Reisert, W1JR popularized design discussed under Diplexer Topics on this web site. The user inputs an IF and presses the Calculate button to have the capacitor and inductor values given in pF and uH respectively. The diplexer schematic is included in the application. Note that the this is for the Q = 1 version of the Bridge-Tee Diplexer.



Style: GUI, File size: 49K, zipped, 22K.

Current Version is: 1 / 19 / 1999

Download the diplexer.zip file

HF Dipole

A very basic program for calculating the length of each leg of a 1/2 wave wire dipole antenna. Program good for 1 - 500 MHz, although intended for MF - HF useage. This app does nothing more than the standard 468/freq (MHz) type calculations. It was written for DOS many years ago and ported to Windows. The output shows the 1/2 wavelength and 1/4 wavelength design wire length in feet and meters. This app is probably of no help to experienced antenna designers.

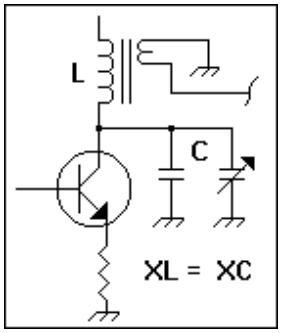
Style: GUI, File size: 46K, zipped, 22K.

Update : Minor improvements made Feb 9, 1999

Current Version is: 2 / 9 / 1999

Download the hf_dipole.zip file

Resonator



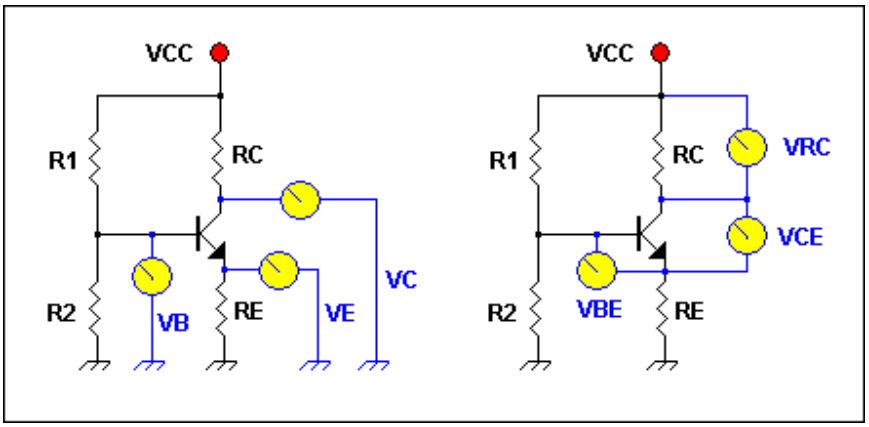
This application calculates the inductor and capacitor values for the tank circuit of a simple bipolar transistor RF amp. The basic schematic is shown above. Enter the center frequency plus the inductive/capacitive reactance you desire and press the Calculate button to calculate the necessary inductance and capacitance for L and C respectively.

Style: GUI, File size: 50K, zipped, 21K.

Current Version is: 1 / 23 / 1999

Download the resonator.zip file

NPN DC-BIAS



This application calculates the various voltages and currents of a simple voltage divider bias NPN bipolar transistor amp. The following is calculated: IB, IC, IE, VE, VB, VC, VCE and detection of Saturation or Cutoff. The user can alter the VCC, VBE, transistor beta and any of four resistor values R1, R2, RC and RE by picking the transistor value from a standard-value resistor table or manually entering the value. The schematic illustrates some of the voltage measuring points on the transistor schematic. *This app is in final BETA.*

Style: GUI, File size: 73K, zipped, 32K.

Current Version is: 16 / 04 / 1999

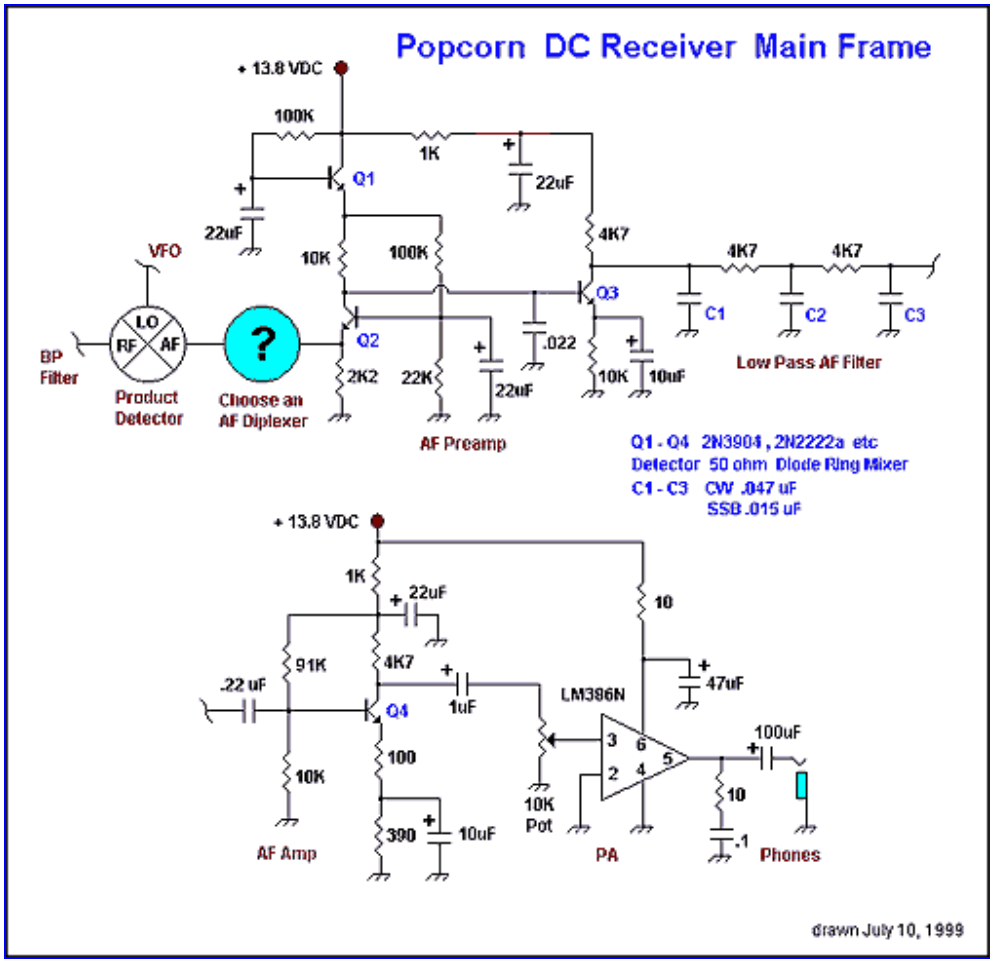
Download the nbias.zip file

Popcorn Direct Conversion Main Frame

Discussion:

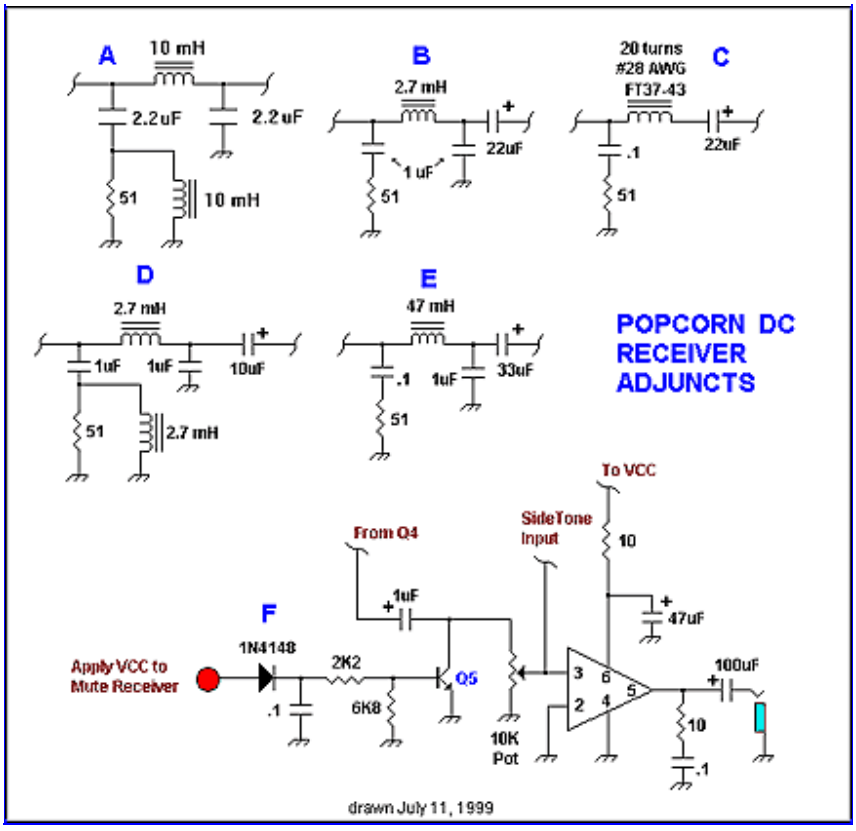
Note: if you click on a schematic, a larger version will appear in a new web browser window.

Shown to the right is the schematic to a low cost popcorn direct conversion receiver main frame. To complete the receiver, a front end band pass filter and a VFO with an output power of 7 dBm is required. This is indeed a frugal project using 4 cheap transistors, an RC low pass filter and an LM386N for output power to a pair of low-impedance headphones. The builder also has a choice of 5 diplexers and an optional mute circuit. This receiver is easily built using Ugly Construction and can be built in 3-4 hours with a bit of luck.



Product Detector and Diplexers

The 50 ohm diode ring product detector can be commercial units such as the Mini-Circuits SBL1 or TUF-1 or homebrewed 50 ohm impedance units. Five simple "diplexers" are shown in the lower "Adjuncts" schematic for you to choose from. The one you choose will depend on available parts, cost and your requirements in a popcorn receiver such as this. These diplexers are mostly of the low pass filter variety and provide a ~50 ohm termination to the diode ring mixer and some matching to preserve the product detector dynamic range. I realize that except for (A) and (D) these audio frequency filters are not truly diplexers and will not provide DC to daylight matching. The intent of this web site is not high performance-high cost



design and please do not confuse it as such. Note that electrolytic capacitors that bypass to ground such as the 1 uF caps must be non-polarized or bipolar for best results.

The (A) diplexer is by W7ZOI and is described on the Diplexer Web Page on this site.

The (B) and (D) diplexers are my designs and the (D) diplexer is the (B) diplexer with out the high pass component.

The (B) diplexer shown has a 3000 hertz 2 pole high pass/2 pole low pass design. This 2nd order filter provides reasonable overall matching Capacitors are standard-value, non-polar electrolytic types.

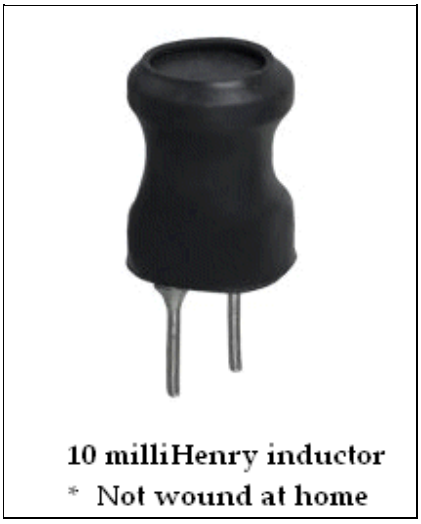
The (C) diplexer is a very basic, but very practical choice for this receiver.

The (E) diplexer is one that I used in one of my first DC receivers and the 47 millihenry inductor is a standard value unit sold by Mouser Electronics and others.

Another diplexer choice for this receiver might be the unit described by Rick Campbell, KK7B in his Binaural I-Q receiver project published in the March 1999 issue of QST.

Update May 15, 2009

There was confusion regarding the 2.7 to 47 mH inductors mentioned on this web page. I originally wound just the 2.7 mH inductor on a ferrite, but not the others. This is not a great idea as losses are high. For millihenries-value inductors, commercial parts should be purchased. A good brand to consider might be Epcos. Sorry for causing confusion.



AF Preamps

The AF preamp section follows that of the Ugly Weekender Receiver designed by Wes Hayward, W7ZOI. I tried many other configurations

and came to the conclusion that these two simple but elegant stages give a winning combination of low noise, good gain, low parts count, low hum and good AM broadcast band rejection. The Q1 transistor decouples the receiver preamp very well and no hum was detected in the headphones providing a well filtered DC power supply was used. The Q2 grounded base amp provides a low impedance termination of the product detector and diplexer stages. Q2 and Q3 are direct coupled and provide lots of gain to drive the succeeding low pass filter without it adding a huge abundance of noise to the signal. The bypass capacitor (0.022 uF) is essential to bypass any broadcast AM detected in the Q1 stage to ground. Other values of capacitors maybe tried, but do not omit this critical part.

Low Pass Filter

I cannot handle listening to a DC receiver on a crowded band without some low pass filtering. The high pitch heterodynes effect my concentration and give me a headache. Nevertheless, it is neat to temporarily listen to an unfiltered DC receiver; to hear the pure and wonderful signals possible by beating RF directly into audio. I prefer low pass to band pass filters at audio and have used many combinations of active filters using discrete components and op-amps, as well as passive designs using AF inductors to build wave filters. This receiver uses none of these devices, however they could be easily substituted for the filter shown. Connected to Q3 is a simple, cheap RC low pass filter based upon the design criteria given on the Discrete Component RC Audio Filters web page on this web site. The cutoff values you calculate will be ballpark and values of 0.047 uF for CW and 0.015 uF for SSB were chosen, but other values could just as easily been used and please do not hesitate to experiment with the caps and/or the resistors to suit the parts you have on hand. For the capacitors in the low pass filters, avoid using ceramic disk type caps if you want the best possible performance. Polyester, polypropylene, polystyrene or polyester film type are all suitable, however, ceramic caps will work if you are really going junk box/low cost.

I attempted to make a wave file to demonstrate the low pass filter. I came right off the headphone jack into the input of the of my 16-bit PC sound card via a step up audio transformer and the results were a little disappointing. Sixty-cycle hum and distortion of loud stations were added by the sound card. The sound file is big (636 KB) and is a digital recording of me tuning through a 30 meter pile-up using the lowest sample rate and frequency possible on my computer. The low sample rate/frequency also degraded the sound somewhat as well, but I decided to put it on the page, warts and all. The DX station was a VK2 and sure did cause a lot of excitement on 30 meters that night around sunset on the left coast. Actually the wave file demonstrates how good the receiver sensitivity and AM radio immunity is. In addition, the low receiver background noise is also very apparent underneath the constant 60 cycle hum. The 60 cycle hum and clipping of loud CW signals is not heard in the headphones and is a soundcard manifestation. Perhaps the best method would be to come of Q4 and go right into the sound card with a smaller line-in signal voltage. Download the popdc wave file

---

AF Driver and Final Amp

Connected to the input and output of the Q4 stage are small value capacitors to provide some high pass filtering for the receiver amplifier chain. Some emitter degeneration is used on Q4 to provide a better termination of the preceding RC low pass filter. The receiver amplifier chain has a lot of gain and when the 10K pot is turned to minimal resistance (cranked) , the LM386N can be driven into distortion. You may want to limit the maximum gain with a series resistor connected to the 10K pot after building and testing this receiver.

The final AF amp is the perennial LM386N, a low cost, easy to use AF amp. Turn it upside down and solder pins 2 and 4 right to your copper ground plane to anchor this part. It can easily be configured to drive a small speaker.

An optional mute circuit is shown in the "Adjuncts" schematic and is labeled (F). This circuit is a simple transistor switch which grounds the output from Q4 and mutes the receiver audio. This circuit switches rapidly and there are no annoying pops or clicks to be heard in the headphones when it is switched. Apply the VCC to the diode as shown to mute the receiver during transmit if the receiver is used in conjunction with a transmitter. Q5 in the mute circuit can be a 2N3904 or 2N2222a or substitute.

In addition, a suggested side tone input to the LM386N is shown. I have started to use simple one section RC filters on the output of my side tone oscillators to smooth the waveform into a more pleasing audio tone.

Conclusion:

This popcorn receiver can be made very inexpensively and has good sensitivity and a reasonable noise level and selectivity. I tested this main frame on 30 and 40 meters and really enjoyed it. This receiver main frame could be combined with an inexpensive VFO using tuning diodes to keep cost down and the popcorn factor up. Although it does not use tuning diodes, a 40 Meter band VFO schematic has been placed on the VFO page.

Here is a YouTube Link using the receiver with a different front end filter and VFO. This is not my radio or video.

A blog post from Peter AK6L --- it's good to see builders moving beyond kits.

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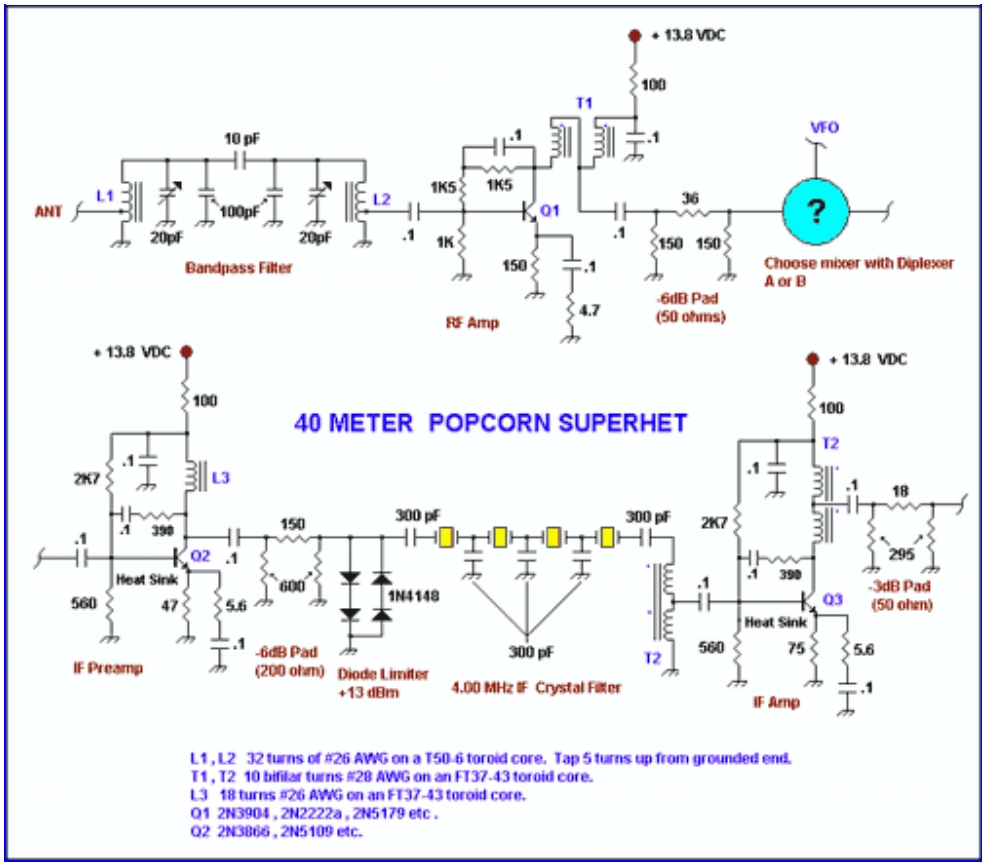
40 Meter Popcorn Superhet Receiver

Discussion:

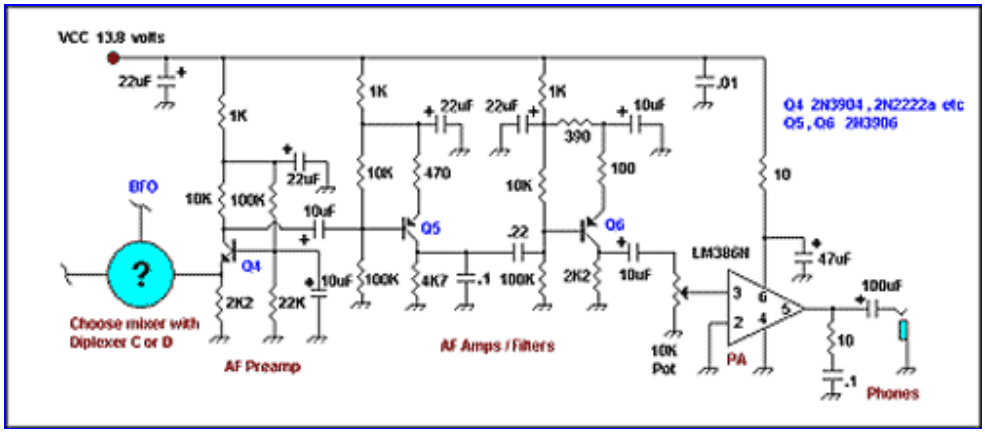
Note: if you click on a schematic, a larger version will appear in a new web browser window.

To the right is the schematic for a no-frills, relatively low-cost CW superhet receiver with a 4.00 MHz Intermediate frequency. There is no AGC or RF gain control, however this receiver has good large signal handling capability. This receiver uses just 6 bipolar transistors and an op amp for reasonable volume into headphones. Much of the ideas/design of the various stages must be credited to Wes Hayward as I borrowed heavily from his previous work and through ideas obtained by discussion. If one were to homebrew the diode ring mixers, indeed this would be a very low cost receiver giving reasonable performance which outperforms any NE602 based superhet receivers that I have built or listened to. Below the main schematic is a diplexer diagram that allows the builder to choose from one of two RF and AF diplexers used to terminate the diode ring mixers.

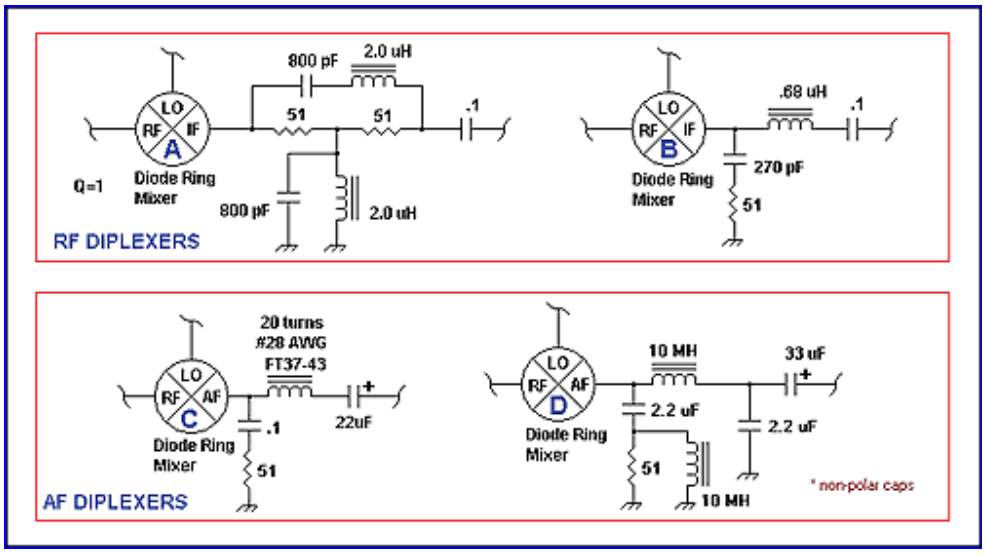


Band pass Filter and RF Preamp

From the 50 ohm receiver antenna jack, first off is a double-tuned band pass filter which was designed by Rick Campbell, KK7B and works very well. The trimmer caps can be the 5 - 20 pF units sold by Digi-key and Mouser. The fixed-value caps in my prototype were inexpensive monolithic ceramic capacitors purchased from Digi-key. Rick used an NP0 ceramic for the 10 pF coupling cap plus silver-mica type for the 100 pF caps in his original design. For possible lower insertion loss, the probable best/cheapest way would be to use all NPO ceramics for the fixed value caps in this filter.



The RF amp is my favorite popcorn RF amp ; a 50 ohm feedback amp. A grounded-gate JFET amp was tried in its place and was also found to be quite suitable and does not require the -6dB pi attenuator that follows the feedback amp as shown in the schematic. The feedback amp's 50 ohm input impedance properly terminates the band pass filter. The -6 dB attenuator pad following the amp to helps provide a 50 ohm input impedance for the mixer and to reduce stage gain which aids in preserving the signal to noise ratio of the receiver. If a builder wants a little more sensitivity, the pi attenuation pad could be reduced to -3 dB however this may effect the receiver dynamic range. The transformer T1 is one of 2 broadband transmission line transformers in this receiver. It transforms the 200 ohm collector impedance to 50 ohms for the succeeding stage.



Mixer and Diplexer

A 50 ohm diode ring mixer (7dBm) such as the Mini Circuits SBL-1 or TUF-1 or homebrew are all suitable. Following the mixer is an RF diplexer of your choice. The more complex Brifge-Tee (Q = 1)diplexer (A) is an excellent design, however maybe overkill in a popcorn superhet such as this. For the (A) diplexer, to get the necessary 800 pF for the capacitors, simply parallel a 470 with a 330 pF or a 120 pF with a 680 pf capacitor. The inductors at 2.0 uH are wound on powdered-iron torroids. You can use # 26 AWG wire and it requires 22 turns on a T37-2 core or 20 turns on a T50-2 core. In addition, you can use a #6 material torroid to wind the inductors. This diplexer is described elsewhere on this web site.

The simpler (B) diplexer uses a ~3 times the IF frequency that I have seen this basic design in many textbooks and articles and provides reasonable matching with a 50 ohm inductive and capacitive reactance. The cutoff frequency chosen was 11.78 MHz as this allows the use of a standard value capacitor (270pF). To wind the 0.68 uH inductor use 13 turns on a T37-2 torroid or 12 turns on a T50-2 powdered iron torroid core. You can easily use 24 - 26 AWG wire for the inductor.

IF Preamp , Crystal Filter and IF Amplifier

Except for the inductors, the IF preamp and IF amp are identical and both warrant a small clip-on heat sink as they draw reasonable current. The standing current maybe increased or reduced by changing the 47 and 75 ohm resistors connected to the Q2 and Q3 emitter respectively. Factors such as available power supply current versus dynamic range requirements may come into play. One may want to stand more current in the IF preamp and less current in the IF amp. For example, the 75 ohm resistor on the Q3 emitter could be increased considerably and/or the 5.6 ohm degeneration resistor could be increased as well if less stage current draw is wanted. The 2N3866 transistor is usually a cheaper way to go for these amps than the 2N5109, but the choice is up to you as you may have something available in your junk box. The 200 ohm -6dB pad following the IF preamp should not be omitted as it helps prevents the stage from seeing reactance's created ahead by the crystal filter. The four diodes form a 13dB limiter to protect the crystal filter should a catastrophically large signal be present in the receiver's front end. They maybe omitted. A -3dB 50 ohm resistive pad terminates the IF amp and helps establish a 50 ohm input impedance for the product detector ahead. **Click here** for more on the IF preamp.

This receiver has a narrow IF Cohn Crystal filter. Bandwidth is ~ 405 hertz, which unfortunately makes tuning quite sharp however this filter is very nice for crowded band conditions. The IF filter crystals should be closely matched in frequency to prevent unwanted ripple in the pass band. Generally, you have to buy 10 and then if you have a frequency counter, use the receiver BFO stage to test your crystals for matching. Pick the closest 4 crystals and use them in your filter. It does not matter if the crystals have series or 20 pF load capacitance, but it does matter that they are matched in frequency within 40 hertz of one another or better for this receiver. For my prototype receiver, I purchased ten 20 pF load capacitance 4 MHz crystals and luckily found 4 that matched each other within 9 hertz! For those builders who do not have a frequency counter, some QRP parts retailers sell matched sets of crystals. It is important to note that the BFO should be set on the high side of the IF frequency as simple crystal ladder filters have a steeper upper passband than lower pass band.

The crystal filter is terminated by the 4:1 transmission line transformer and then 50 ohm impedance of the IF amplifier. The -3dB pad following the IF feedback amplifier helps to terminate the crystal filter by helping ensure a 50 ohm IF amp input impedance and should not be omitted. Place a 75 and a 220 ohm resistor in series to get the required 295 ohm resistance on each leg of the pi attenuator.

Many may balk at just one stage of IF amplification, but since there is no AGC and this is a CW receiver, it works well. A feedback amp is once again used to provide correct input and output impedances for stages connected to the IF amp. Following the IF amp is another attenuator set for -3dB and then a 50 ohm diode ring mixer.

Product Detector, AF Diplexer and Audio Amplifiers

The mixer/detector can be SBL-1 or TUF-1 types or homebrew if you want to reduce costs further as the mixers are the single most expensive components in this receiver.

Again a choice of diplexers is required. The (C) AF diplexer is very simplistic but very practical if you are trying to keep costs low. The (D) diplexer is designed by W7ZOI and is from the Diplexer Web Page on this site.

Following the diplexer, a grounded base audio amp provides a 50 ohm termination to the product detector. AF gain and some AF filtering are provided by Q5 and Q6 which together attenuate frequencies less than 72 hertz and greater than 638 hertz. This amplifier pair are described on the discrete AF filters web page on this web site. Keep your leads short on all the AF transistors.

The final AF amp is the perennial LM386N, a low cost, easy to use AF amp. Turn it upside down and solder pins 2 and 4 right to your copper ground plane to anchor this part. There are a number of low-noise alternatives to the LM386 available which are generally more expensive but would be quite suitable. Discrete component AF amps can also be used, but a popcorn part such as the LM386N maybe cheaper and easier.

VFO and BFO

VFO schematic

BFO schematic

Alternate version of this receiver

Construction Ideas

When constructing any project, build in small modules and test each one separately. For instance, the AF amp should be built first and then tested by injecting a very low-level audio frequency tone into that stage and listening for output in your headphones. Every QRP workbench should have a simple AF tone oscillator from a schematic similar to the ones used for keying side tones in CW transmitters. The encased oscillator should have to a 100K or so potentiometer connected to the output to vary the output signal amplitude. Generally use maximum resistance on the 100K pot to start with and reduce this resistance slowly as the in-test amplified oscillator output could be very loud!. After testing the AF amp, build the 3rd AF preamp stage including the 10K panel mounted pot so you can vary the gain going into the AF amp. Now inject the AF oscillator output into the input on the pot and vary the 10K pot to ensure that the stage you built is working. It should be a lot louder now and should go up and down in volume with the 10K pot. Finally build the remaining preamp stage and once again test the circuit with your AF oscillator. The output into the phones should be painfully loud now when cranked up! The next stage to build would be the BFO. If you do not have a scope, peak the tuned circuit by watching the S meter on a radio receiver located nearby. Ensure that you put a load on the output winding of the BFO such as 47 ohm resistor to ground. A small piece of wire can be used as an antenna if the BFO signal is too weak to activate the S meter on your receiver. Once peaked, you can now use the BFO to match your IF crystals. To use the BFO to match your crystals, use a small wire to bypass or disconnect the 60 pF variable capacitor that is used to connect the crystal to ground. In other words, the bottom lead of the crystal is connected to ground with a short piece of wire. This makes testing your crystals a little more scientific as the variable capacitor cannot influence the crystal frequency during testing. You can also use the BFO in conjunction with a scope or voltage probe to test the various RF amps in the receiver. I do this all the time with my scope. Proceed with this build a stage, test a stage method and you should be rewarded with a functional end product.

KK4RF's version of the 40 Meter Band Popcorn Superhet

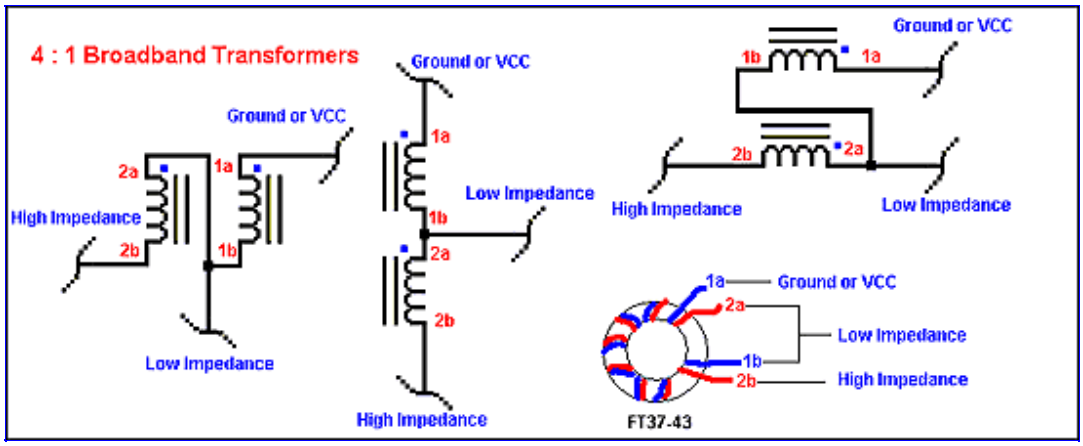
Marty, KK4RF emailed me and described his version of the popcorn superhet from this web page and contributed some great info and photos of his receiver. Of note is Marty's use of Radio Shack IC boards for mounting the components for each stage other than the VFO circuit. This is yet another variation from ugly construction that I have also used which works extremely well. Marty built the VFO using pure ugly construction and I was glad to hear that he is enjoying good frequency stability even with the lid off the VFO enclosure.

He built the receiver into an old Heathkit HW-12 single-bander case from the 1960's and it is a very attractive receiver to say the least. He found an old National Velvet Vernier Drive at the Virginia Beach Hamfest this year and used it to tune the VFO. Don't you love Hamfests! For the BFO he used an BFO circuit with a 4 mHz crystal from a different receiver project (a project that never quite worked.) He built a small power supply and located it along with the BFO under the chassis.

I like Marty's generous use of ground plane and neat stage layout. He reports good selectivity with his IF filter and apparently built four superhets that did not work before building this popcorn version. This is more a testimony to Marty's perservance to home building than to this receiver design in my opinion. I won't tell you how many rig failures I have personally incurred, as it would take a long time! Many thanks to Marty, KK4RF for the feedback and great pictures.

Broadband Transformers plus Diode Ring Mixers

Discussion



There are 2 basic types of broadband transformers used in most QRP work — conventional and transmission line style. Both types may be wound on ferrite toroids, pot cores or rods, however, I only discuss the toroidal transformers we employ to give a 4:1 impedance transformation.

I use these transmission line transformers on many projects on the QRP / SWL HomeBuilder web site. For MF and HF uses, a ferrite core permeability of 850-900 is generally required and the FT37-43 ferrite core proves suitable. Shown above are 3 equivalent schematics of the 4:1 transmission line transformer. You'll probably find that the center drawing easiest to conceptualize, however, with closer examination, all 3 schematics are the same and transform signals from unbalanced 50 Ω impedance up to 200 Ω unbalanced impedance or visa-versa.

The high impedance is 200 Ω and the low impedance is 50 Ω in all cases. It is important to know that these transformers are symmetrical and the points labeled Ground or VCC can be switched with the point labeled High Impedance. *Click on the schematic to enlarge it.*

Winding the 4:1 Transformers

Wire Twisting

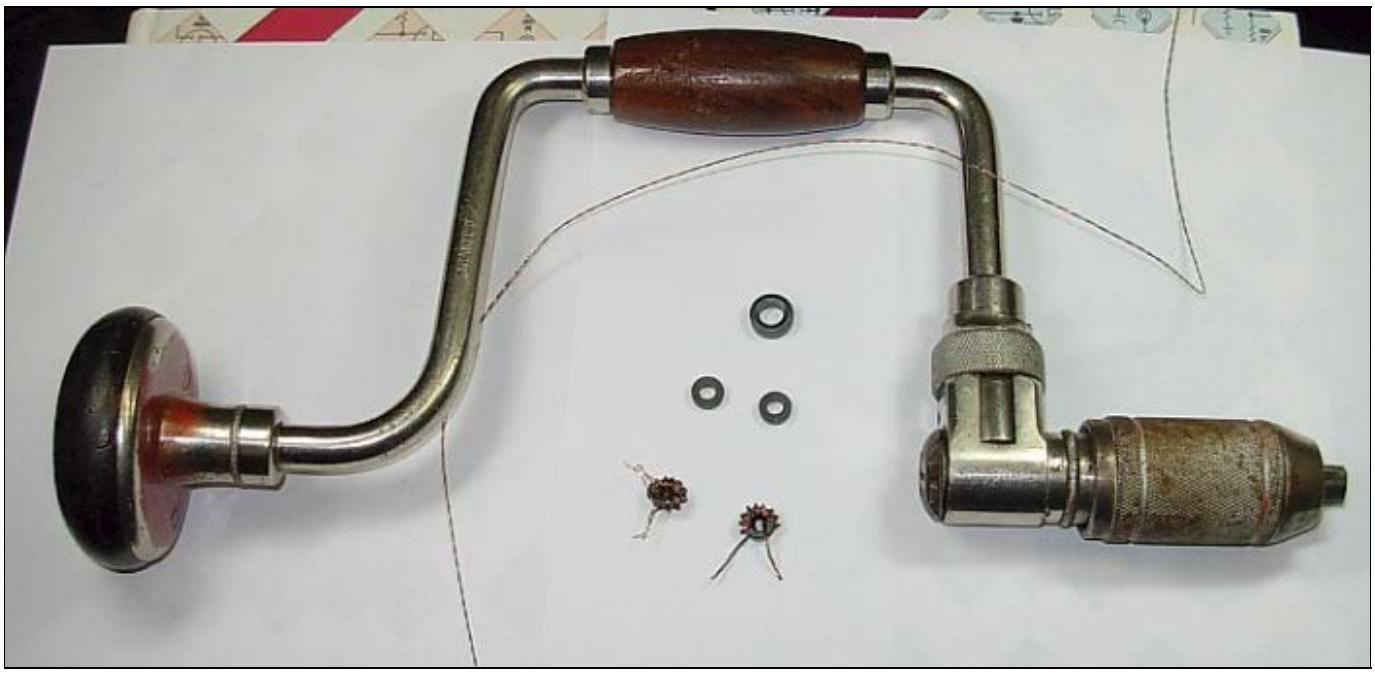
Transmission line transformers are wound with bifilar (2 wires — generally twisted together). Winding these transformers is very easy. All you need are two ~18 cm (7 inch) pieces of #28 AWG enamel coated wire and an FT37-43 ferrite toroidal core.

A shop vise, a ruler, plus a brace and bit hand drill may aid your construction — I bought my brace and bit drill at a garage sale for 2 dollars. You need to twist the 2 pieces of wire together to get ~3-4 twists per cm (8-10 twists per inch) in the wire. To do this, loosely twist the wires at one end and place these twisted ends in a bench vise. Next, place the free wire ends together in your brace and bit drill chuck (no drill bit) and tighten up the chuck so that the wires are held securely.

Try to keep wire length and tension equal. Start turning your drill to twist the wires together and every once and a while measure how many twists per cm with a ruler. When you get to 3-4 twists per cm (8-10 twists per inch) you're done and then may trim the excess leads with a wire cutter in preparation for final winding and soldering.

At VHF, I often use just 3 or 4 total turns on an FT23-43 toroid with a piece of wire just a few cm long. I place the 2 wires in the vise and twist them using a pair of pliers held parallel to the wire.

Transmission line transformers will also work if the wires are untwisted. 3-4 twists per cm only serves as a non-criticial guide. Never wind your simple 4:1 transformers with bare wire.



A brace and bit hand drill plus a vise provides a good way to twist your wires.

Final Wiring and Soldering

Leaving a 2.5 cm (1 inch) lead, wind ten complete loops through the toroidal core leaving a small gap between the start and finish leads.

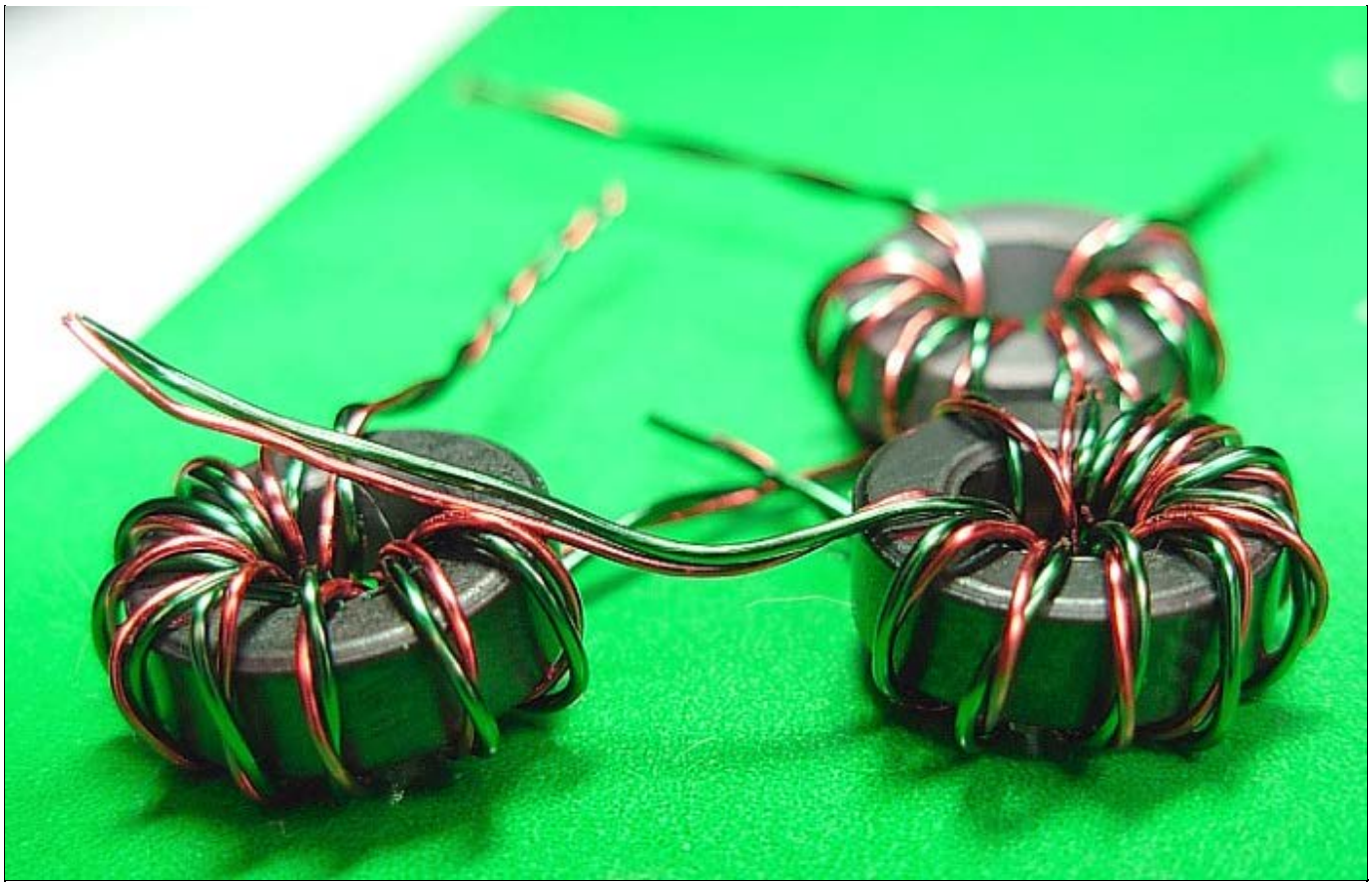
Untwist the leads a little so that you have 4 separate wires. One set of these wires wires will be called winding #1 and the other winding #2. You need to identify them and further break them into 1a, 1b and 2a and 2b. Generally I regard the the top two windings as (A) and the the bottom two wires (B), however, use whatever system works best for you. Strip off the enamel at the tips of all four leads and then get your ohmmeter or better yet, a beeping continuity tester.

Start on one of the top (A) wires by connecting the ohmmeter or continuity beeper to it and then touch one of the bottom wires and then the other bottom wire. Whatever bottom wire (B) shows continuity with your top wire should be marked along with the source top (A) wire with paint, liquid paper, tape, or whatever you like.

I prefer to wind 2 different colored wires if possible.

Designate the marked wire pair winding number 1. You may also want to test for a short circuit — there should be no connection between wire set 1 and wire set 2 at all. So now you have 2 wires sets, winding set 1 is marked and winding set 2 is unmarked. The top two wires are arbitrarily labeled A and the bottom two wires are labeled B . Refer to the schematic above for clarification. Connect 1b to 2a and twist them together and then solder. Your done!

It's really easy to make these things don't you think?.



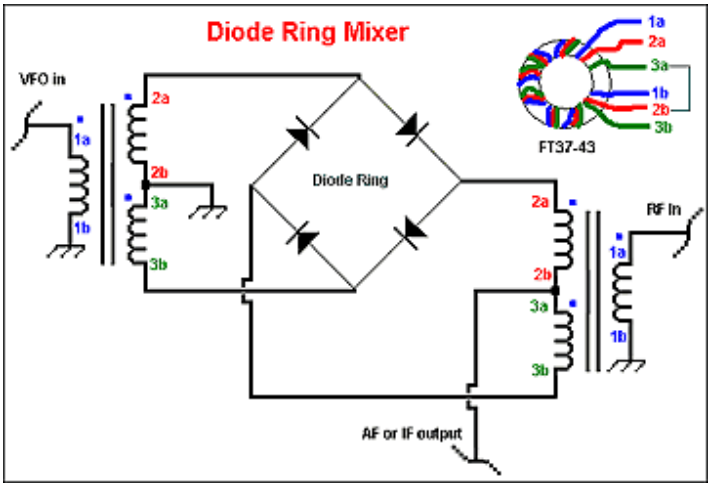
A trio of bifilar transformers wound on FT37-43 ferrite toroids. 2 colors of wire reduces errors and speeds up construction. Consider making up 5 at a time, so you have them on hand and do not have to interrupt your experiments.

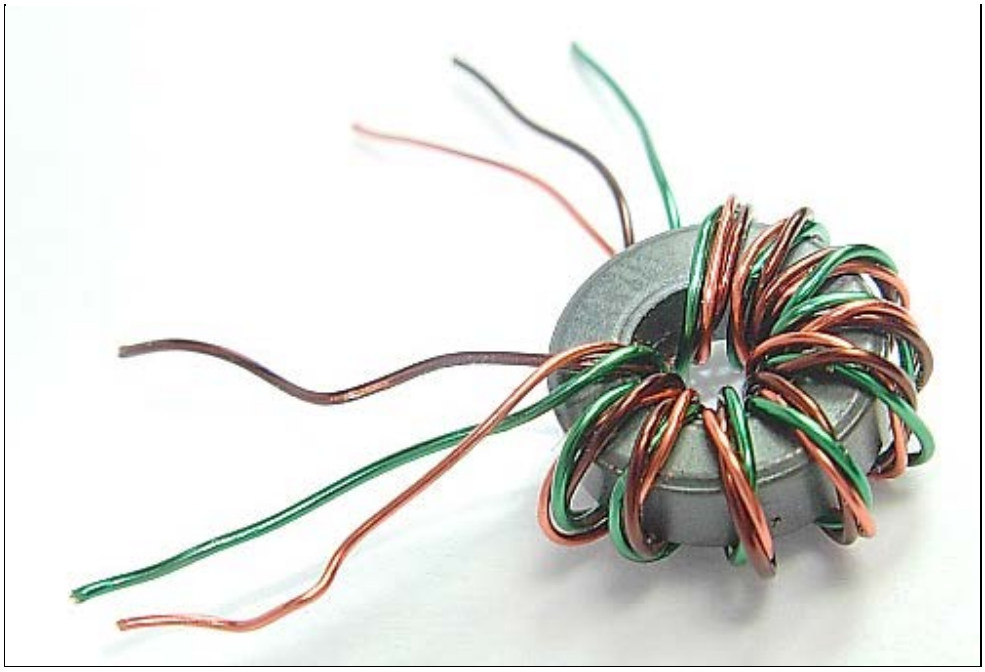
Homebuilding Diode Ring Mixers

Discussion:

Easy to make, homebuilt diode ring mixers give a low-cost alternative to commercial diode ring mixers. A double-balanced diode ring mixer has 2 unbalanced to balanced transformers and a diode ring. The impedances at the three ports is 50 Ω. The transformers are wound with #28 AWG enamel coated wire on a FT37-43 ferrite toroidal core using a trifilar (3 wire) technique.

The wire twisting and winding technique is done as described above for the bifilar transformers. The connections 2b and 3a are twisted together and soldered. Again you will have to develop a technique to help you distinguish the wires from 1 another. *Click on the schematic to enlarge*

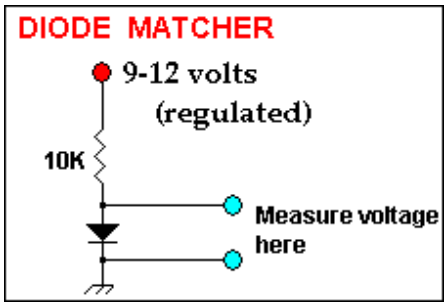




A trifilar transformer wound using 3 colors of wire on an FT37-43 toroid

A sample of the enamel coated copper wire collection I wind inductors and transformers with. In stock are wire gauges from 18 to 34. Like everything else, this collection started small and grew over time. Be vigilant for bargains and when you find a good price, purchase a whole bunch as it does not go bad. The Belden wire (orange spool ends) is over 40 years old and the enamel insulation remains perfect.

Diode Matching for Mixers



Discussion

For optimal results Schottky or Hot-Carrier diodes should be used. However, common diodes such as the 1N914, 1N4148 or 1N4454 are all quite suitable and are much cheaper. The four ring diodes should be matched to help mixer balance and thus carrier suppression. At MF and HF the most critical matching required is the forward voltage drop across the diode and this is easily performed with a sensitive voltmeter.

Set your voltmeter on the 2 volt scale to give you 3 decimal places for matching the voltage drops. Try and find 4 diodes close to one another. In addition, best results maybe obtained if all the diodes are the same type (i.e. all 1N4148) and if they are all from the same manufacturer. Look above for easy schematic to match your diodes with a voltmeter. Give the diode under test at least 20 seconds to warm up and stabilize before taking your voltage measurement.

Diplexers Topics

Introduction

My original web page on diplexers was rather incomplete and received some criticism from electronic engineers albeit the focus of this web site is "popcorn" designs. Wes Hayward, W7ZOI provided me some excellent schematics, analysis and simulations for diplexers which terminate doubly balanced mixers and these are presented below. After this section, the W1JR Bridge-Tee RF Diplexer from the original QRP HomeBuilder diplexer web page is presented along with new commentary and simulations by W7ZOI.

The final section presents a practical diplexer for terminating a product detector. All graphical images labeled as Figures 1-24 are copyright and property of W7ZOI and may not be presented elsewhere. Updated September 23, 2000.

W7ZOI Diplexer Notes

The usual amplifier is a two-port circuit. That is, it has an input port consisting of two terminals and an output consisting of two more. One terminal (ground) can be shared between the ports. Many filters are also two-port networks, including most of the ladders networks we use so often. Many other networks have three or even more ports. A common example is a mixer, which has three ports. Another example of a three port network is a diplexer. This linear network is usually designed around two port filters where one end of two different filters are paralleled to form an input port. This is illustrated as Figure 1. The purpose of a diplexer is usually to force a frequency constant impedance to occur at the input port, even though we usually only use one of the two output ports for signals. The simplest form of diplexer uses a pair of 1 element filters, a low pass and a high pass. This is shown in Figure 2.

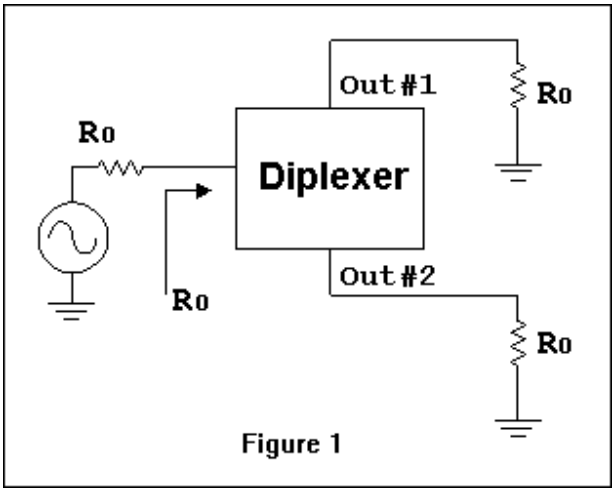


Figure 1

Where the

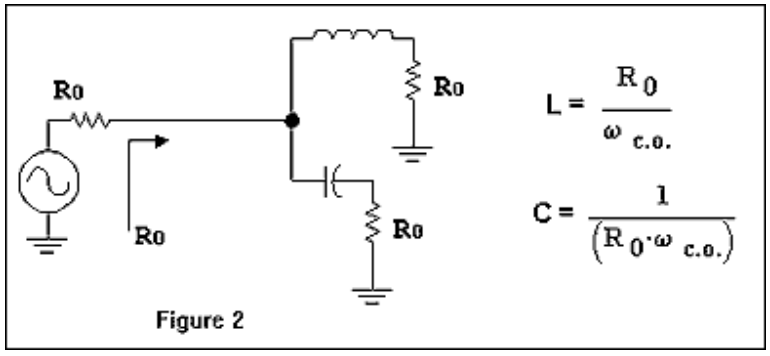


Figure 2

equations give the L and C that provide a perfect match. The angular frequency is called the cross-over. A familiar example is the cross over used in audio systems. The network that splits signals is a diplexer. Here is an example where both outputs are used. Another form of diplexer is the band pass/band-stop combination. This is shown in Figure 3:

Let's now consider further some examples, some that work and some that don't work as well. First, let's look at an audio diplexer that follows a product detector in a DC receiver. The load of interest is the first audio stage, which has a 50 Ohm input resistance. The diplexer offered is Figure 4. Note that this is not the combination of filters. It just looks like a low pass with an extra resistor. The response of this circuit is shown in Figure 5. The transmitted signal never gets

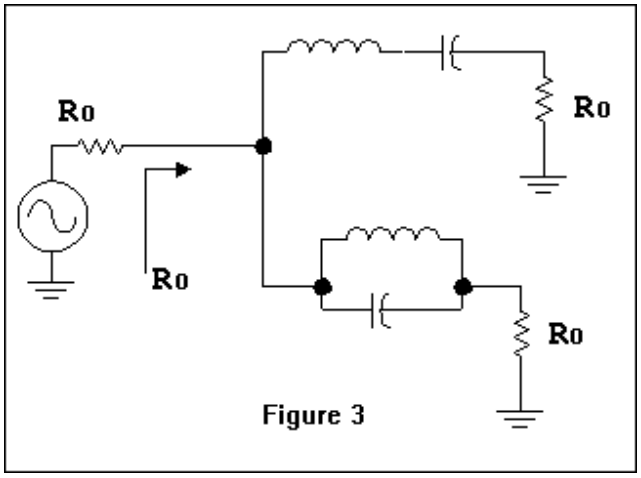


Figure 3

up to the desired 1 volt in the low pass passband while the impedance match, represented by reflection coefficient, never gets down to the desired zero. The response is just that of a lossy low pass filter.

In 2012; Click on many of the diplexer images to see the original sized version

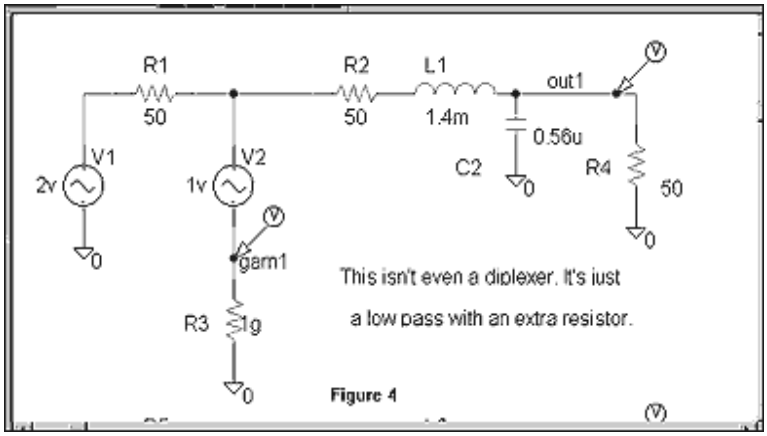


Figure 4

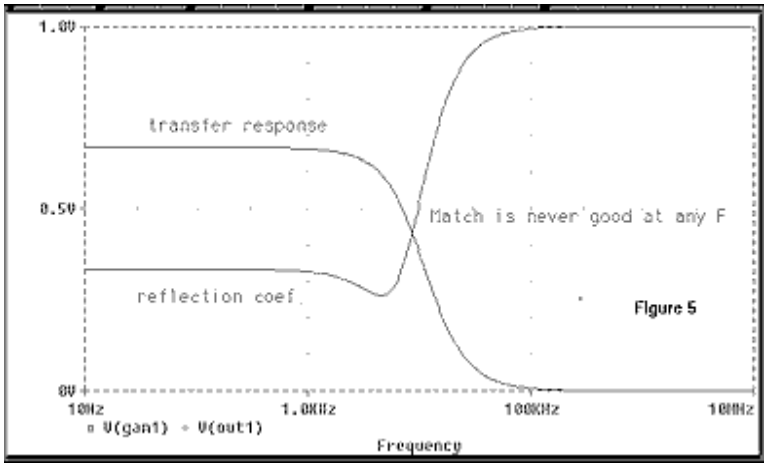


Figure 5

The normal filter circuit without the extra resistor is Figure 6. The corresponding output response is shown in Figure 7. Note that the transmitted signal is now up at 1 while the reflection is down to zero, both within the passband. Transmission goes to zero while reflection is 1 in the stopband.

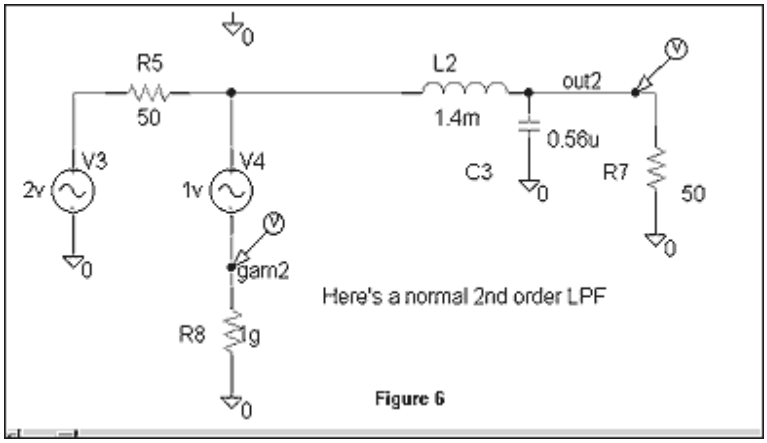


Figure 6

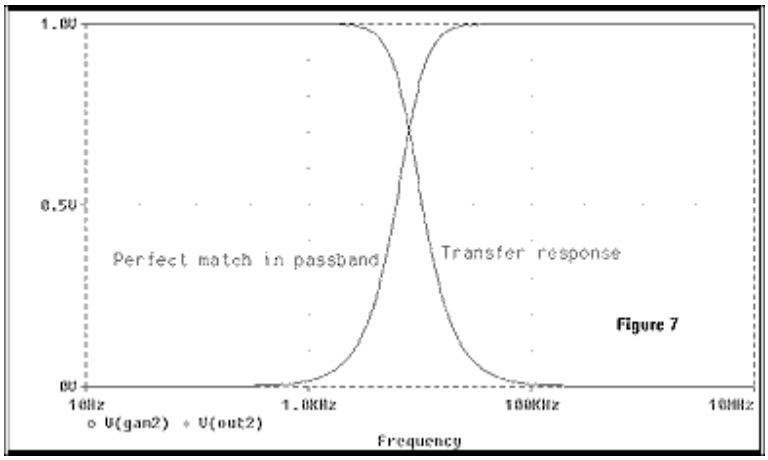
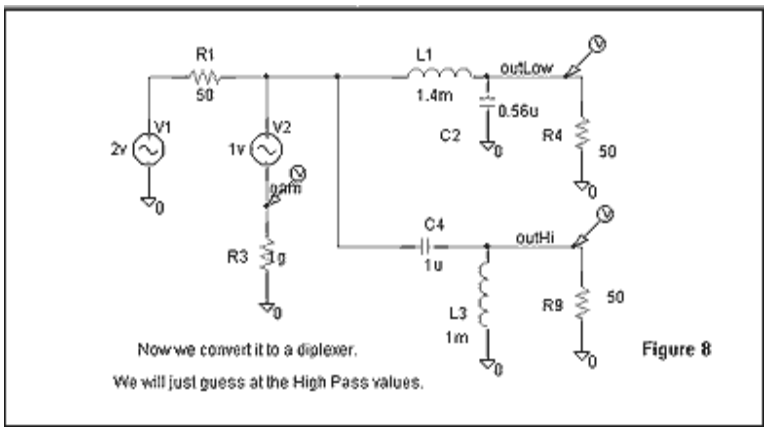
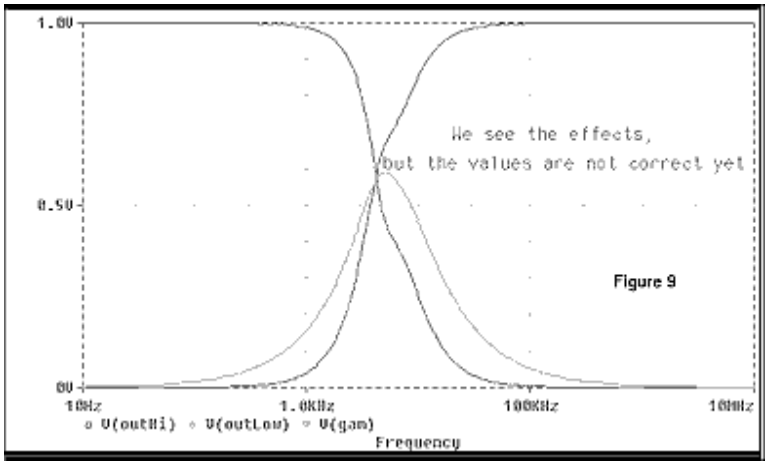


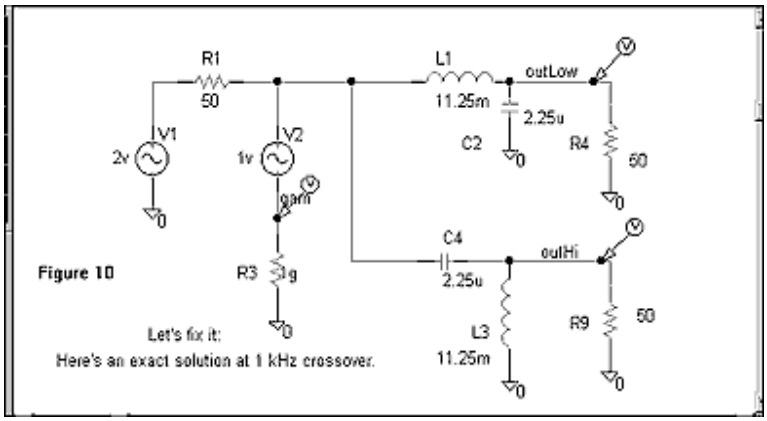
Figure 7

Now let's use the low pass and put a high pass with it to try to form a diplexer. This is shown in Figure 8 where we now have just guessed at component values. The response, shown in Figure 9, has high pass and low pass outputs that we might expect. The match is good at the frequency extremes, but is only so-so in the transition band.





Let's now look at a carefully designed pair of two element filters. The circuit is Figure 10 an is a final example. The corresponding response is Figure 11. It is hard to see, for the response merges in with the baseline. However, the reflection is zero and it is zero everywhere. This filter was designed for a 1 kHz crossover, so it can be scaled to other frequencies with ease.



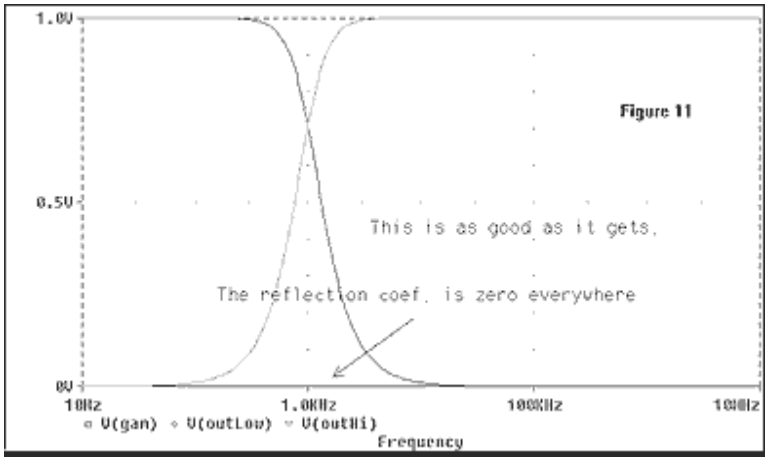
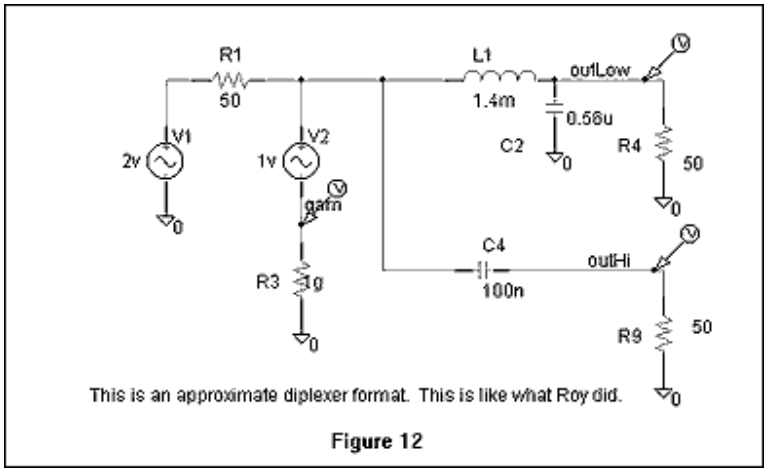
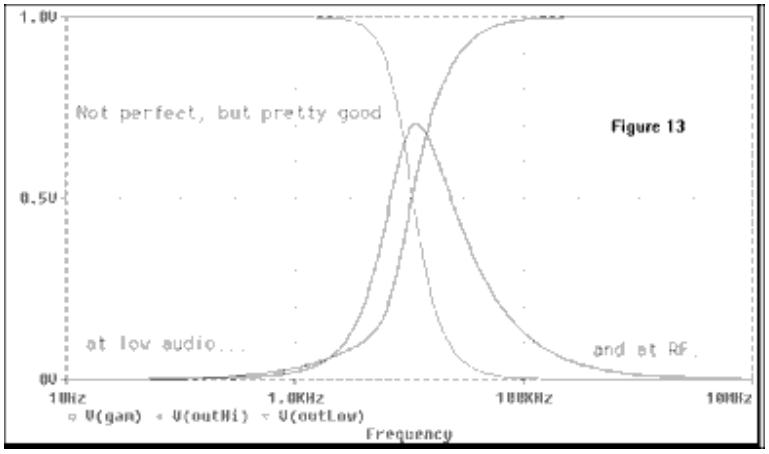


Figure 12 is another final audio example. This circuit is very similar to the one used in the past by Roy Lewallen, W7EL, although the inductor was smaller at 100 uH in his Optimized rig. The response of this diplexer is shown in Figure 13. This is not perfect, but it is probably quite a good performer in typical receiver situations.





Finally, here's a higher frequency example. 5th order low pass and high pass filters are combined. The filters have a cross over at about 150 MHz. Note that there is a slight reflection in the transition band. This is probably just the result of our having rounded some values in the design process. Figures 14 and 15. An outstanding reference on this is Nic Hamilton, G4TXG, "Improving Direct Conversion Receiver Design," Radio

Communications, April 1991.

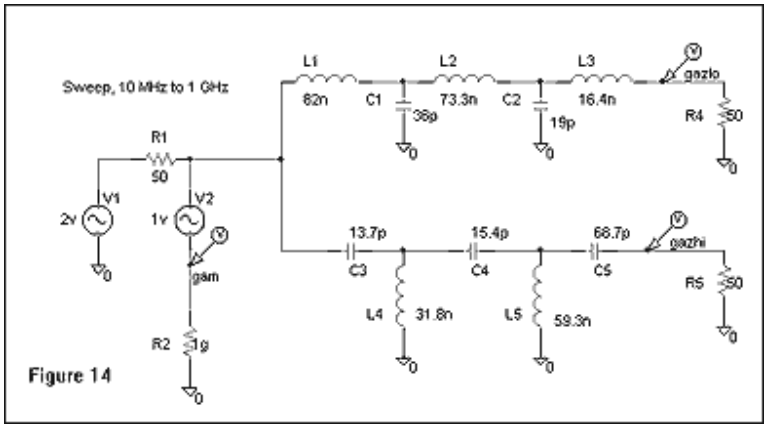


Figure 14

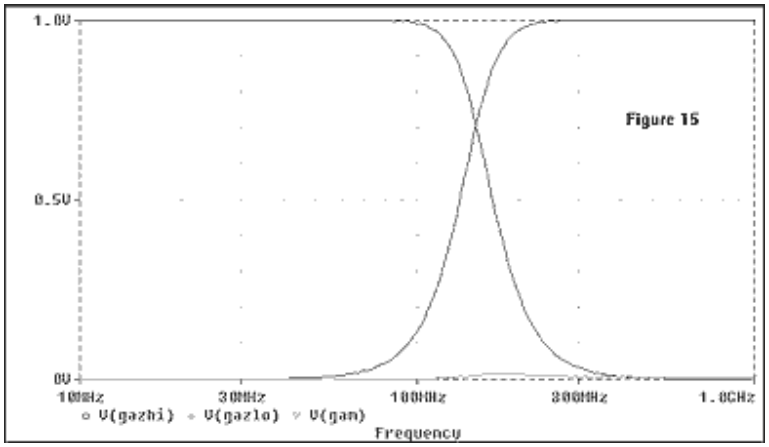
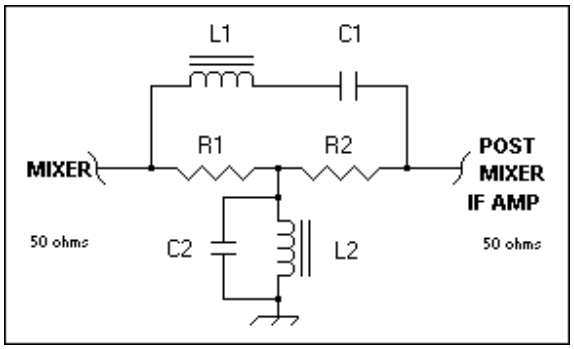


Figure 15

Bridge-Tee RF Diplexer



This is an excellent bandstop/bandpass diplexer popularized by Joe Reisert W1JR. This easy to build diplexer has a low parts count and is easily built using Ugly Construction. Resistors R1 and R2 present a 50 ohm impedance to the mixer output and a 50 ohm impedance to the input of the post mixer amplifier. The IF frequency is passed through the diplexer while out of passband RF is given a low impedance path to ground. The capacitance for C1 is generally built up by substituting the nearest standard value capacitor or by placing 2 or more capacitors in parallel with each other to achieve the desired value. The same procedure is then repeated for the C2 capacitance. For more strenuous purposes, a portion of C1 and C2 or the inductors L1 and L2 can be variable and adjusted on the bench. The inductors can easily wound on powdered-iron toroid cores. I have used T50-2 or T50-6 type toroids with good results. The Q of the inductors is 1.

It is possible to design a more generalized form of this diplexer with a higher loaded Q in the resonators. The diplexer shown and used in the program has a Q of 1. This was used by W1JR in his VHF/UHF World Column in the now defunct HAM Radio Magazine for March and November 1984. It was also more recently used by Jacob Makhinson, N6NWP in his A High-Dynamic Range MF/HF Receiver Front End in QST for February 1993. The actual formulae for this diplexer is far more complex than the simplified formula shown below or used in the program, but both provide a very good approximation for the Q = 1 version as used by W1JR and N6NWP. If you wanted Q=10, the series tuned circuit would use L that is 10 times as high with C to resonate. The parallel tuned circuit would then use C that was 10 times higher with L to resonate.

A supplemental web page with some hard-core mathematics for this diplexer can be found on the Diplexer Supplemental Page.

Simplified Formulae (Q = 1):

R1 and R2 are always 51 ohm resistors.

Inductors L1 and L2 -> 50 / (6.283 * frequency in Megahertz)

Capacitors C1 and C2 -> 1 / (6.283 * 50 * frequency in Hertz)

Example 1: For a 9 Mhz IF , L1 and L2 = 0.88 microhenrys and C1 and C2 = 350 picofarads

Example 2: For a 4.92 MHz IF , L1 and L2 = 1.62 microhenries and C1 and C2 = 647 picofarads

I wrote a simple program to do the math for the Q = 1 version. Download the Bridge-Tee RF Diplexer Diplexer Program

---

Comments and analysis by W7ZOI

This is a double ended version of the first order bandpass/bandstop design presented earlier. But it's a good one, within the constraints of what it can do. The first is the simulation schematic for the diplexer, which is better termed a Bridge-Tee Diplexer. (There are bridge Tee filters and attenuators too.) That figure is entitled Figure 16.

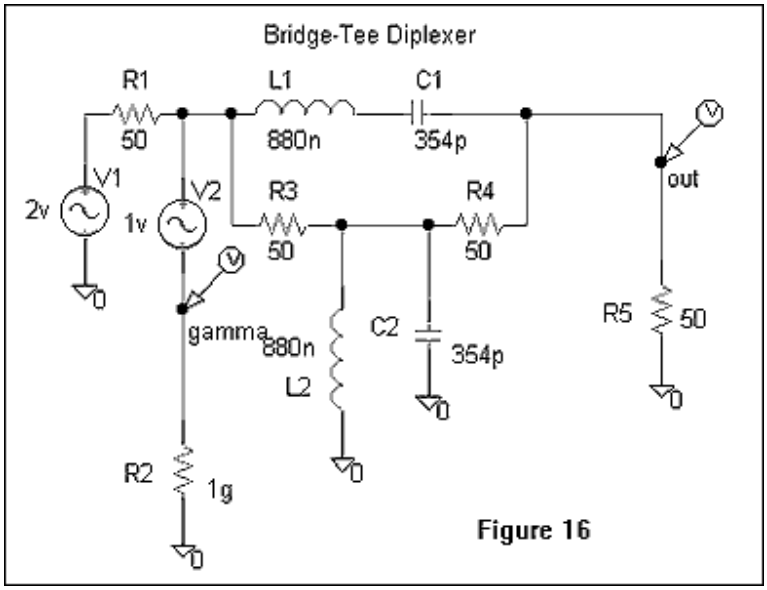


Figure 16

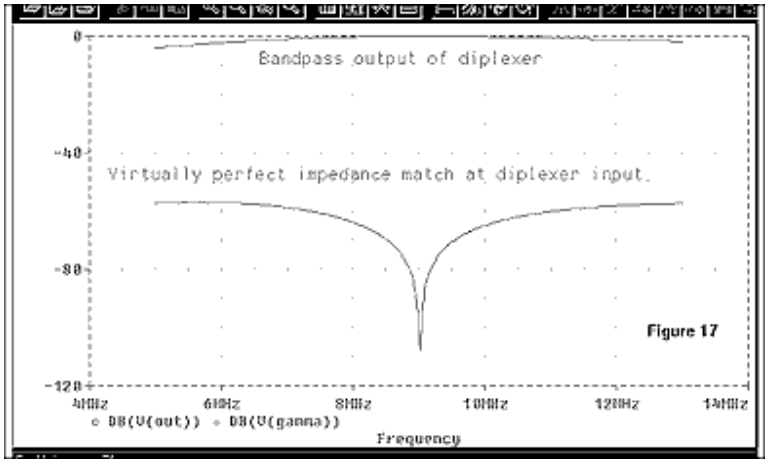


Figure 17

The response for this circuit is shown in Figure 17. This is extremely good. The through response is very flat owing to the low Q of the series tuned circuit. But even better is the match. It is very good. Indeed, it would have been perfect except for slight roundoff errors that occurred as we designed the networks.

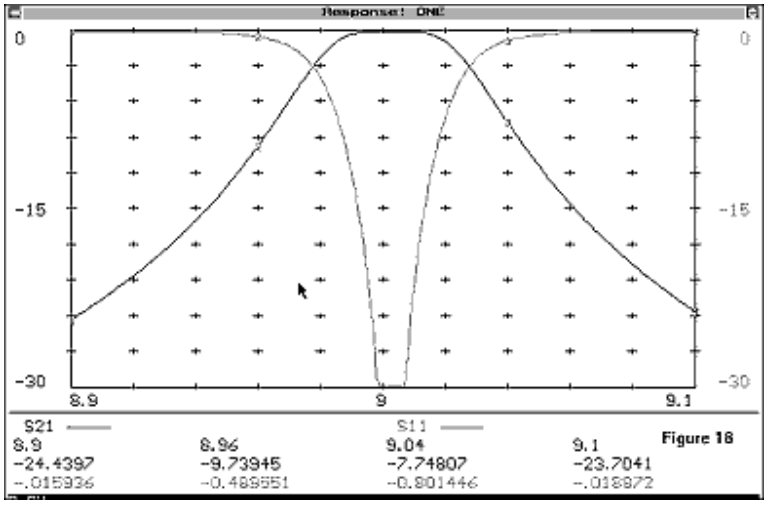


Figure 18

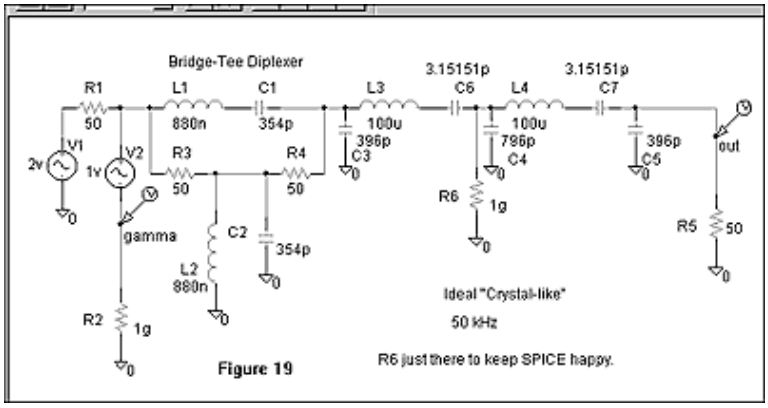
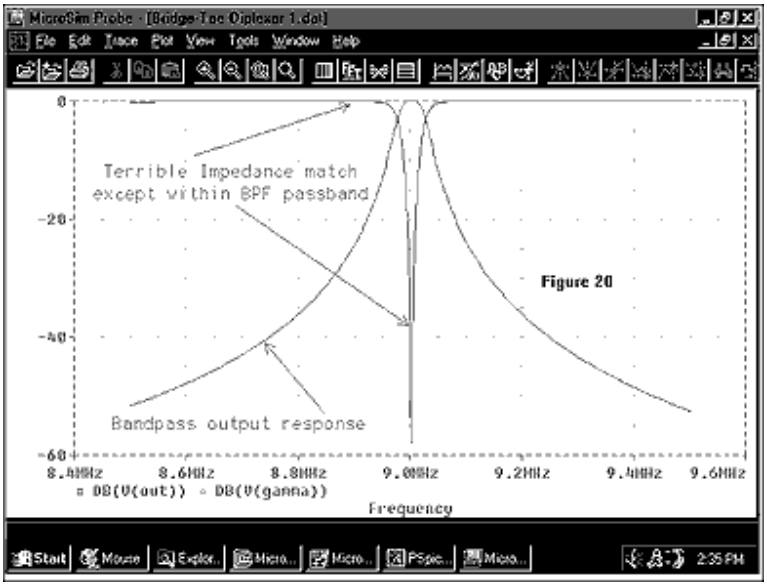


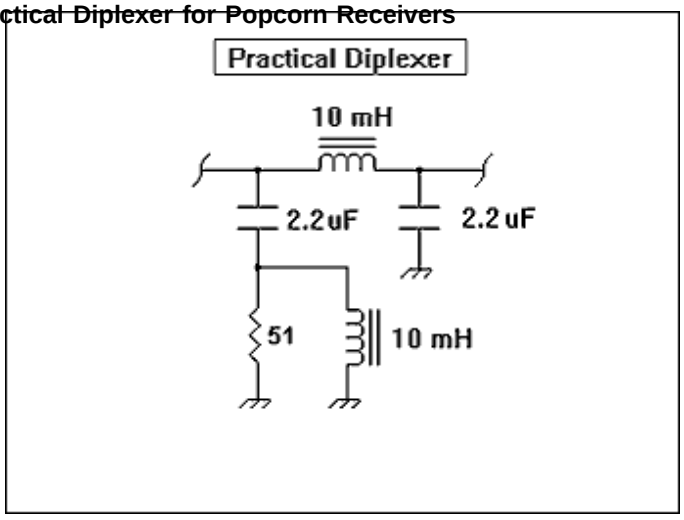
Figure 19

This kind of thing works fine if you really have a perfect match following the diplexer. But what if you don't. There are some places where they do not do the job that some folks think they will do. For example, a diplexer WILL NOT cause the impedance to be flat if it is followed by a filter. The diplexer must still be properly terminated at both output ports. In Todd's usual applications, he is worried about providing a good mixer termination for a product detector. The audio amp that he uses will usually have a common base first stage and that will present a good wideband load to the diplexer, so he is okay. But other folks have placed a diplexer after a switching mode mixer that then drives a narrow filter. The diplexer then does little good. To illustrate this situation, I designed a "crystal like" two pole LC bandpass filter with a 50 kHz bandwidth. This represents the general case where we try to put a diplexer between a mixer and a filter. The filter response by itself is shown in Figure 18. The schematic for the diplexer and following filter is in Figure 19.

The response for the combination is in Figure 20. Here we see a passband response that is fine; it's just the repeat of the filter response we already saw. However, the input impedance looking into the diplexer, the impedance that would be seen by a mixer, is terrible. The return loss is 0 dB at all frequencies except where we get within the passband of the filter.



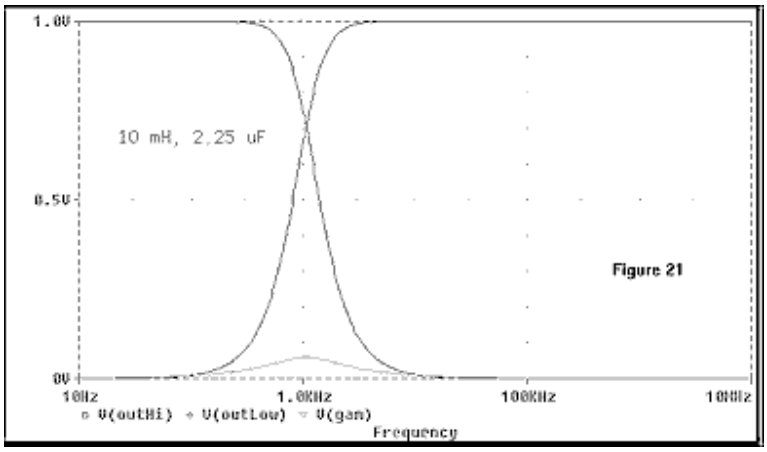
Practical Diplexer for Popcorn Receivers

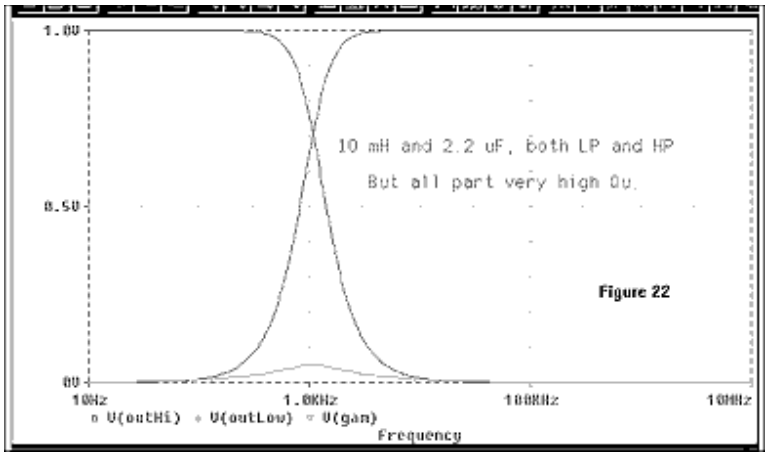


Building a diplexer to follow a product detector is not a cheap endeavor. Audio inductors and capacitors such as metalized polyester film types are not common in many builder's junk boxes. I really like the design shown in Figure 10 and wished to use it because it uses just 2 inductors and capacitors which is in keeping with the popcorn nature of this website. The main difficulty is that the inductors and capacitors are not standard value types and series connecting components to achieve the desired values would add to both the cost and size of the finished product.

Obviously, it will not likely match from DC to daylight. That is not the intention of this simple design or this web site in general. I asked Wes to place just 2 standard value capacitors and inductors in the Figure 10 diplexer design and see what happens.

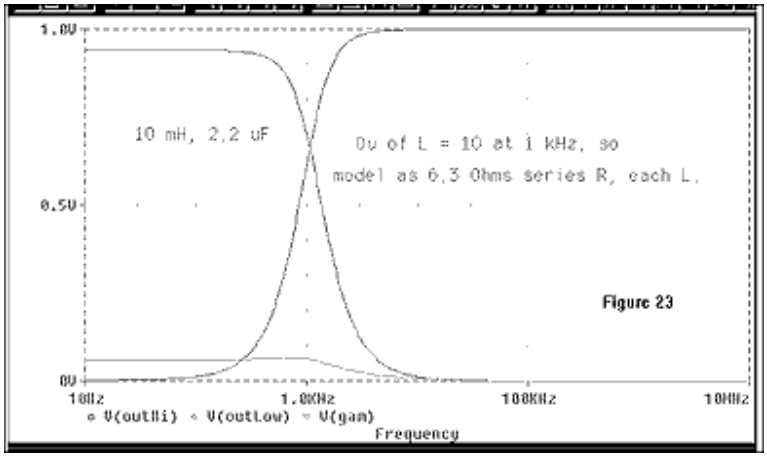
Here was his response to my request:
OK, here are some "practical values." Note that things don't really change that much. We start with 11.x mH and 2.25 uF. Change the inductor to 10 mH and get Figure 21. Then change the cap to 2.2 uF and see almost no change in Figure 22.

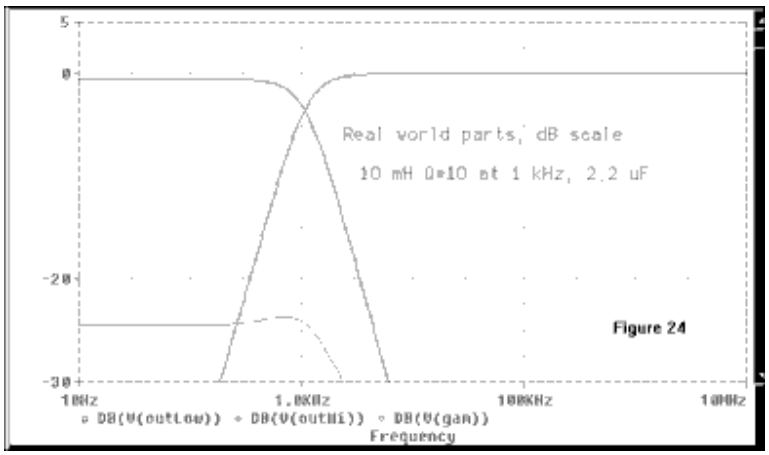




But now move into the world of even greater reality and acknowledge that many of the inductors we use at audio are very low Q. Change Qu of L to 10 at 1 kHz, so I put 6.3 Ohms in series with each L to get Figure 23. And do the same thing, but with a dB scale, for Figure 24. Note that we can see the difficulties, but things are still pretty good. We see some loss (about 1 dB) in the low pass path and less than perfect match. But the match is still very good. 20 dB is about 1.1:1 vswr, much better than 99.9% of the hams can really measure. (A 10 dB match is about 2:1.) Hope this is what you were after....Wes

It was and I will use this "practical" diplexer in my next popcorn DC receiver project. Note that the practical diplexer input and output impedance is 50 ohms and the 2.2 uF caps should not be polarized capacitors such as regular electrolytic types which have a positive and negative polarity.



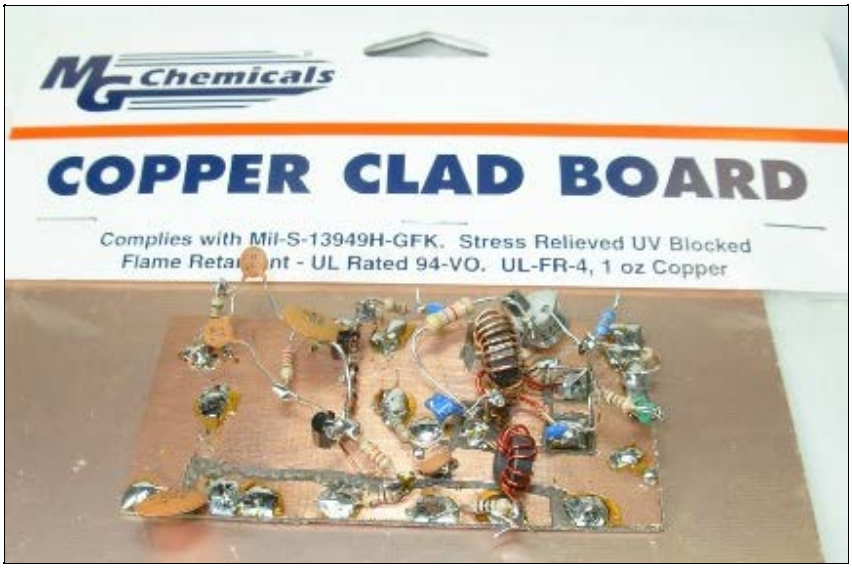


Many thanks to Wes Hayward, W7ZOI for his work on this page.

A version of this web page in Russian Cyrillic

Ugly Construction

Discussion

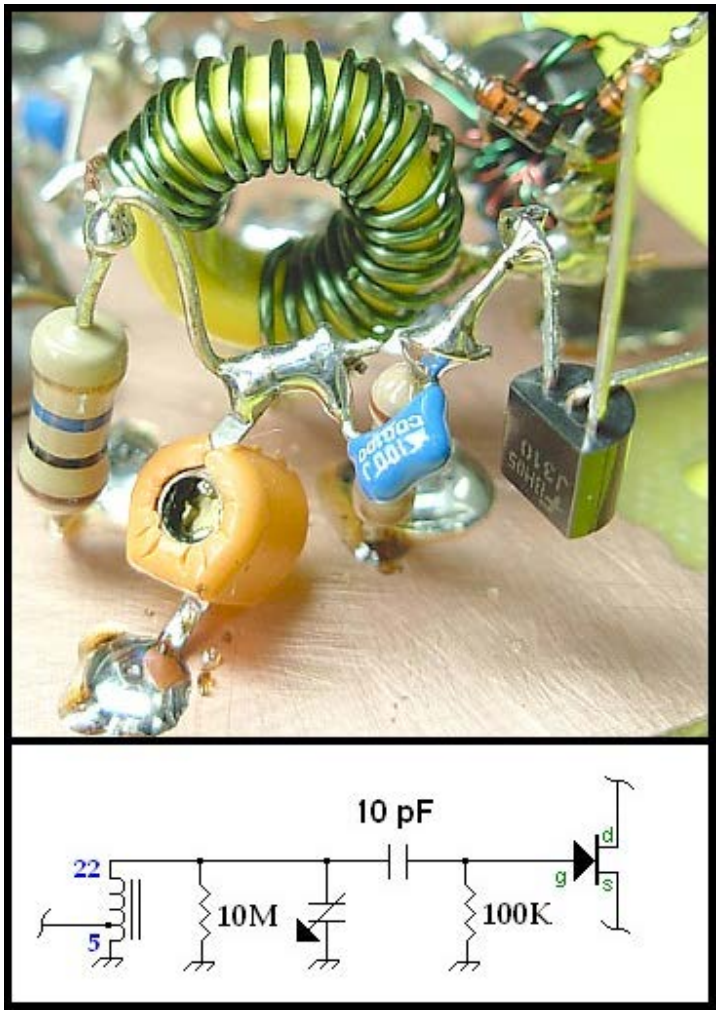


We enjoy many ways to build electronic circuits. For example, you might breadboard on a perforated circuit board, an etched PC board, a sheet of copper clad board, or even a piece of copper wire. In the hay day of tube electronics, builders used terminal strips and point-to-point wiring within the project chassis. I mostly breadboard using Ugly Construction.

Ugly Construction, "dead bug", or "ground-plane construction" involves building circuits on top of a double or single-sided copper clad board (copper side up for single-sided board). The copper ground-plane provides a low impedance ground and mechanically supports the parts soldered to it. Component leads requiring grounding are soldered *directly* to the copper surface, while the ungrounded leads of these parts anchor any ungrounded parts connected to them. Isolated sections called stand-offs hold other ungrounded or remotely located parts.

Example stand-offs include high value resistors (10 Megohm or greater), terminal strips, or small copper islands glued onto or cut into the copper ground-plane. Parts such as transistors, IC's or commercial diode ring mixers are generally flipped upside down and anchored by their grounded lead(s). Metal encased parts such as crystals can be grounded by a short wire or directly soldered upside down to the copper board.

DC voltage wires, or decoupling resistors may be supported by soldering 1 lead of a bypass capacitor to the ground plane while the other lead holds the DC voltage carrying part up off the copper board a short distance. I mostly use grounded caps for stand-offs and not 10M resistors.



Shown above — a 10 Megohm, half-watt, stand-off resistor anchors the "hot" inductor terminal plus supports the ungrounded trimmer capacitor terminal. The coil's 24 gauge wire provides additional mechanical stability. The signal loss from adding the 10M resistor was about 0.1 volts peak to peak in 1 experiment.

Ugly Construction allows the experimenter total control over the design of a project and in my opinion, its greatest strength is *speed*. Ugly Construction yields rapid and flexible bread boarding — very appealing for prodigious home builders.

The Origin of the Term "Ugly Construction"

Roger Hayward, KA7EXM and Wes Hayward, W7ZOI coined the term "Ugly Construction" while writing the "Ugly Weekender" published in the August 1981 issue of QST. I asked Wes about this in 2009. The term was a takeoff from the 1958 book entitled The Ugly American by William Lederer and Eugene Burdick.

A big part of the learning of the QST article was Ugly Construction! The term and the bread boarding technique emphasized the fact that there is no correlation between the "prettiness" of a construction project and the way it works. According to Wes, the goal had a couple of corollaries. "First, people like myself who do NOT have the knack for doing pretty projects can still build successful radios. Second, is that we all need to look at our projects after the fact to discover what it is that really makes them work well. In the case of the Ugly Weekender, the thing that makes it fly is that there is a wonderful ground plane with that PC board material".

Indeed, this transceiver functions very well; especially after you temperature compensate the VFO. Wes also built versions for the 30 and 80 meter bands. I have versions on 15, 40 and 80 meters. The transmitter portion is a true QRP classic; both as a Ham radio transmitter and because it promoted "ugly" ground-plane or dead bug bread boarding techniques to the scratch homebrew community.

Classic Ugly Construction

This term emerged in Spring 2010 and describes the archetype popularized by Roger and Wes Hayward. All ungrounded leads not anchored to other parts are attached to the copper ground-plane via high ohm resistors — *no* glued pads or islands cut in the copper board.

In-situ comparisons of a 10 megohm resistor versus islands cut into the ground plane and glued-on Manhattan-style pads demonstrated that the resistor had the lowest capacitance; around 1 pF versus 4 pF or greater for the pads or islands. Click for a high resolution transmitter chain built

with 100% Classic Ugly Construction. Click for a crystal oscillator.

Above — *Classic* Ugly Construction using a high ohm stand-off resistor. The top of the 10 megohm resistor is the VCC connection point. It feeds a 150 ohm / 47 uF decoupling network connected to a transistor collector resistor. Bypass capacitors also serve as stand-offs and I prefer thick lead (100 volt or greater) caps for stand-off duty.

Above — The original Ugly Weekender. Photograph used by permission of Roger, KA7EXM.

Above — The original Ugly Weekender. Now this is *Classic* Ugly Construction. Please refrain from building this transceiver and adding modern notions such as a PIC microcontroller keyer — that's just wrong!

Photograph used by permission of Roger, KA7EXM.

Ugly Construction Variants

The most popular Ugly Construction variant is called Manhattan style. Manhattan or "paddy board" construction uses small square or round pads cut or stamped from PC board that are glued copper side up onto a large copper clad board also placed copper side up. The small pads or "islands" serve to anchor ungrounded components. Components soldered to the pads such as transistors or ICs are generally not positioned upside down like in Classic Ugly Construction. Many Manhattan style builders use IC sockets as well. These hobbyists sometimes build beautiful looking layouts — Manhattan is a wonderful bread boarding technique. Google for more information. The best Manhattan construction and copper board chassis bashing I've seen comes from Dave, AA7EE. Click for a blog describing his version of the WBR regenerative receiver. *Linked with the permission of Dave, AA7EE.*

Another interesting variant is used by Dick Pattinson, VE7GC. The circuit board is placed copper side up and holes are drilled and countersunk so that the holes are isolated from the ground plane. Ungrounded components are connected underneath the main board through the countersunk holes. There are many such variations. Each Ugly Construction variant has advantages and disadvantages.

On this website, stand-offs are created by cutting a few lines into the copper board with a small, motorized hobbyist cutting tool; with high megohm value resistors, and occasionally by a small Manhattan style pad or 2. Manhattan pads are great for supporting components needing solid anchoring such as a trimmer capacitor or potentiometer.

Classic Ugly Construction dominates circuits breadboarded after May 2010.

The motor tool may also be used to grind off the copper underneath where VFO toroidal inductors will lie, so that the inductor Q is not effected by the being glued onto a copper surface. In audio projects, I may also grind off the copper around the copper board mounting bolts so that they are isolated from the chassis and do not provide multiple grounds and create the potential for ground loops.

Is Ugly Construction Less Reproducible than Manhattan?

I've received well over 1000 emails about Ugly Construction since launching the site in 1998. Some feel that circuits made with Manhattan pads are more reproducible than Classic Ugly Construction builds. This might be true, but to my knowledge nobody has performed a comparison trial.

The important question is why would this happen? I've read/heard opinions that the stray L and C from the long component leads associated with Ugly Construction might wreck circuit reproducibility, but respectfully disagree from DC to HF. In microwave breadboards, we fabricate lumped element inductors and capacitors (*i.e. precisely dimensioned Manhattan pads*) right into the PC board — Manhattan pads glued all around a breadboard may potentially exhibit much more stray L and C than a few component leads in an Ugly build. Also, wise Ugly builders keep their lead lengths short where it counts: for example, RF bypass + ground and at the input/output of a BJT or FET that offers gain into UHF.

I conjecture human error probably inflicts more problems for Ugly Construction builders — Manhattan building, with its slower pace might trigger less mistakes by newbies. Still, too, Manhattan builders tend to make prettier, squared and aligned circuits and it's easier to spot trouble — plus they look nicer in photographs and some builders carefully document and photograph their builds for others to admire and strictly copy. I've see Manhattan build photos where every resistor tolerance band pointed in the same direction — wow! I think it might be difficult to put such a 'work of art' into an RF-tight metal box for much-sought isolation.

Further, in Ugly Construction — upside down parts might wreak havoc on the "spatially challenged" builder. Who knows? I'm comforted knowing that kit sellers who provide a screened printed circuit board with explicit instructions, still must provide major email support to mitigate build errors. To err in an ugly fashion is human?

Whatever variant of construction you choose, it's sure to be a winner!

Further Discussion

Wire

Non-stranded (solid core) copper wire such as the 22 AWG 3-color package sold by Radio Shack seems a good choice for hook up wire. With non-stranded wire, you do not have to worry about little stay wire hairs causing shorts and it's easier to wrap around components leads. I use red for wires that carry positive voltage, green for grounding and black for wires that carry AC signals short distances. In addition, RG-174 or shielded wire is used to carry AC signals for distances greater than 10 cm, and for connecting stages requiring 50 ohm input or output impedances such as diode ring mixers or low-pass filters.

Your Health

Please consider the following safety comments:

For regular soldering, ensure ventilation of your room — flux fumes can be harmful. Open your shop window and/or use a small fan to improve fresh air intake;

Whenever possible, perform high wattage soldering outdoors;

When grinding paths on copper clad boards, wear a small particulate respirator, gloves, plus ear and eye protection and most importantly; **do it outside** for yours and your family's health;

Wash your hands after soldering and handling freshly cut, fiberglass dust laden copper clad boards.

Soldering Irons

For soldering copper clad boards together, AC grounds on tube guitar amp chassis and performing antenna work, I currently use a Weller SP 80L (80 watt) soldering iron. It is heavy and unwieldy, so you have to be very careful when its plugged in. These high wattage soldering irons produce lots of smoke.

Shown above is an 80 watt "heat torch". My main soldering irons are typically in the 30-35 watt range. Consider keeping at least 1 back up soldering iron, as you never know when a soldering iron is going to burn up. My current 35 watt iron is shown below. These Weller irons have a built in lamp which lights when they are plugged in; a very nice feature. I also keep a small stock of new soldering iron tips.

Copper Clad Board

Some builders ask about sources for copper clad board. I personally use boards made by MG Chemicals as they have dealers in my city and are reasonably priced and good quality. Try the search words **copper clad board** plus **your country name** in your favorite web search engine. A few links follow, but as I have only used boards sold locally, I can't comment about the online companies.

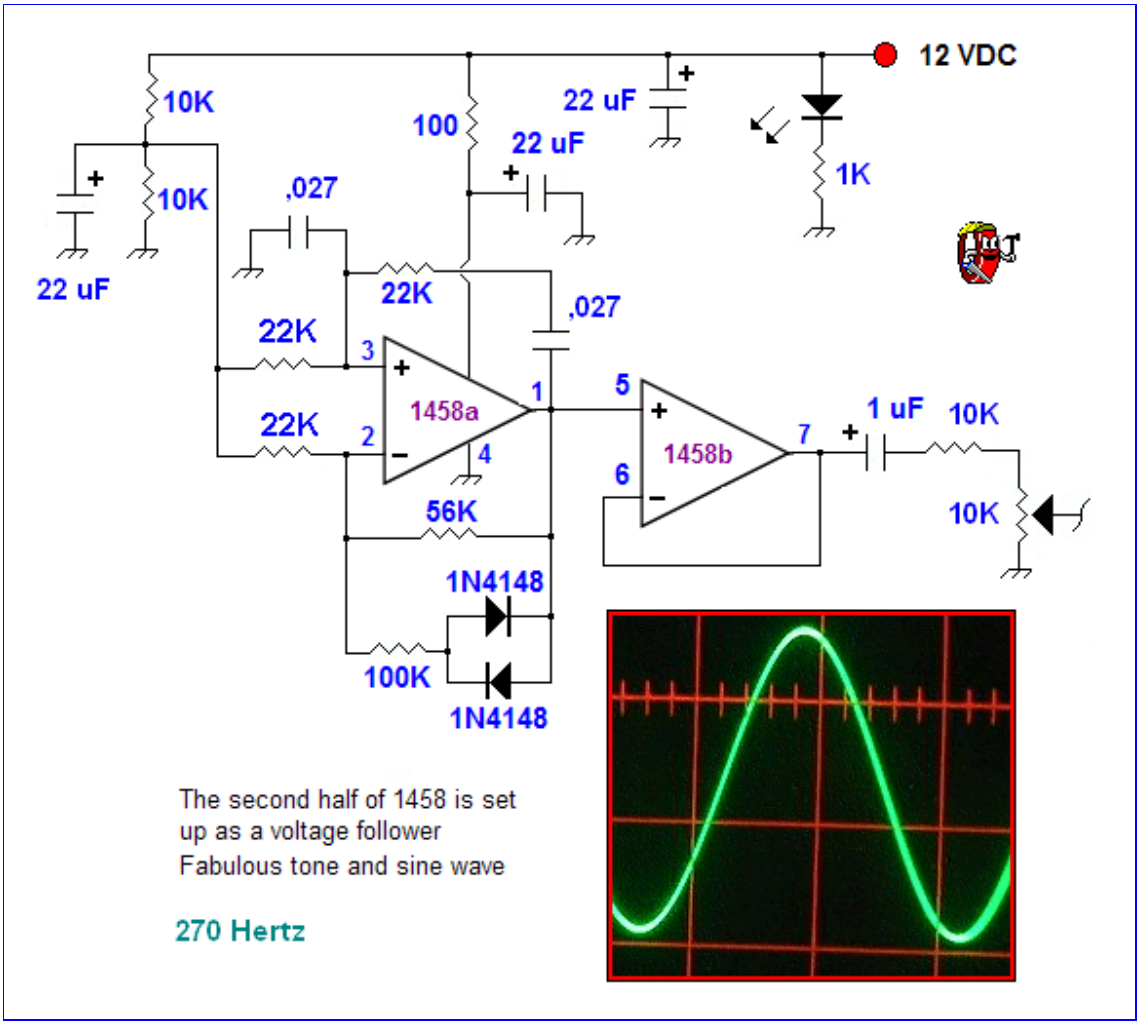
MG Chemicals Worldwide distributer index

Electronic Goldmine Online store

Circuit Specialists Online store

Miscellaneous

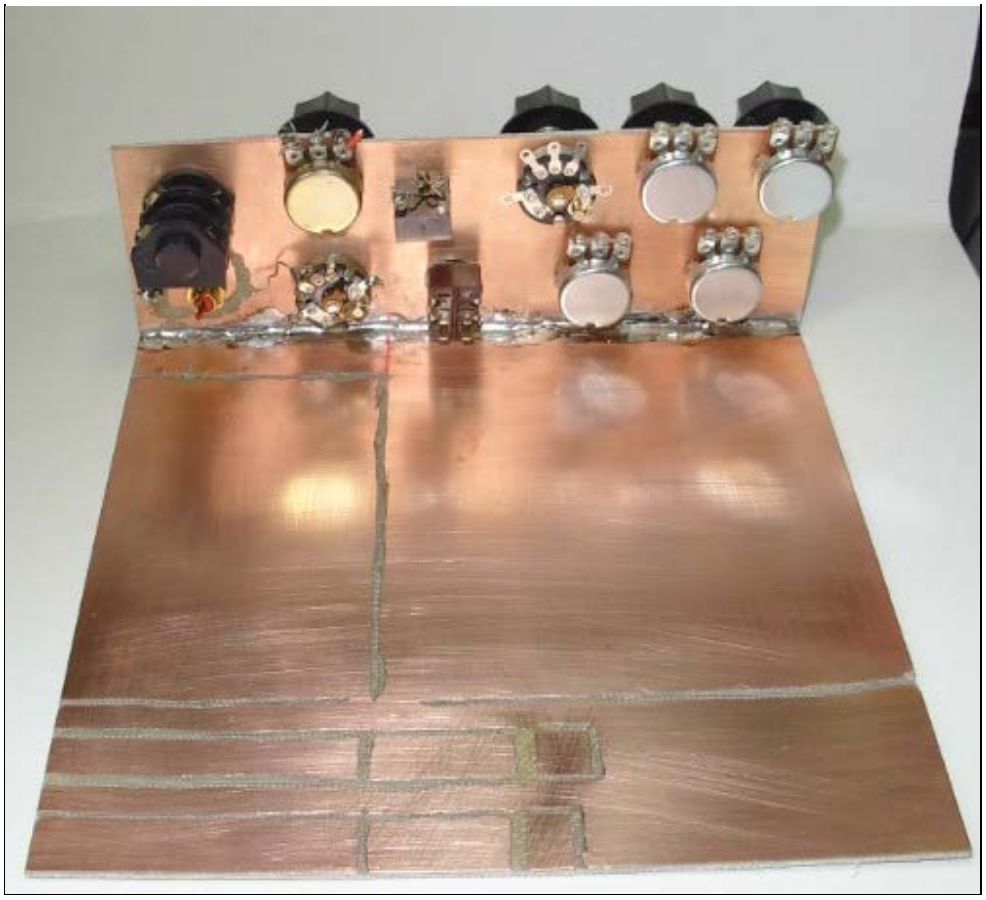
Shown below is a schematic and the Ugly Constructed version of it.



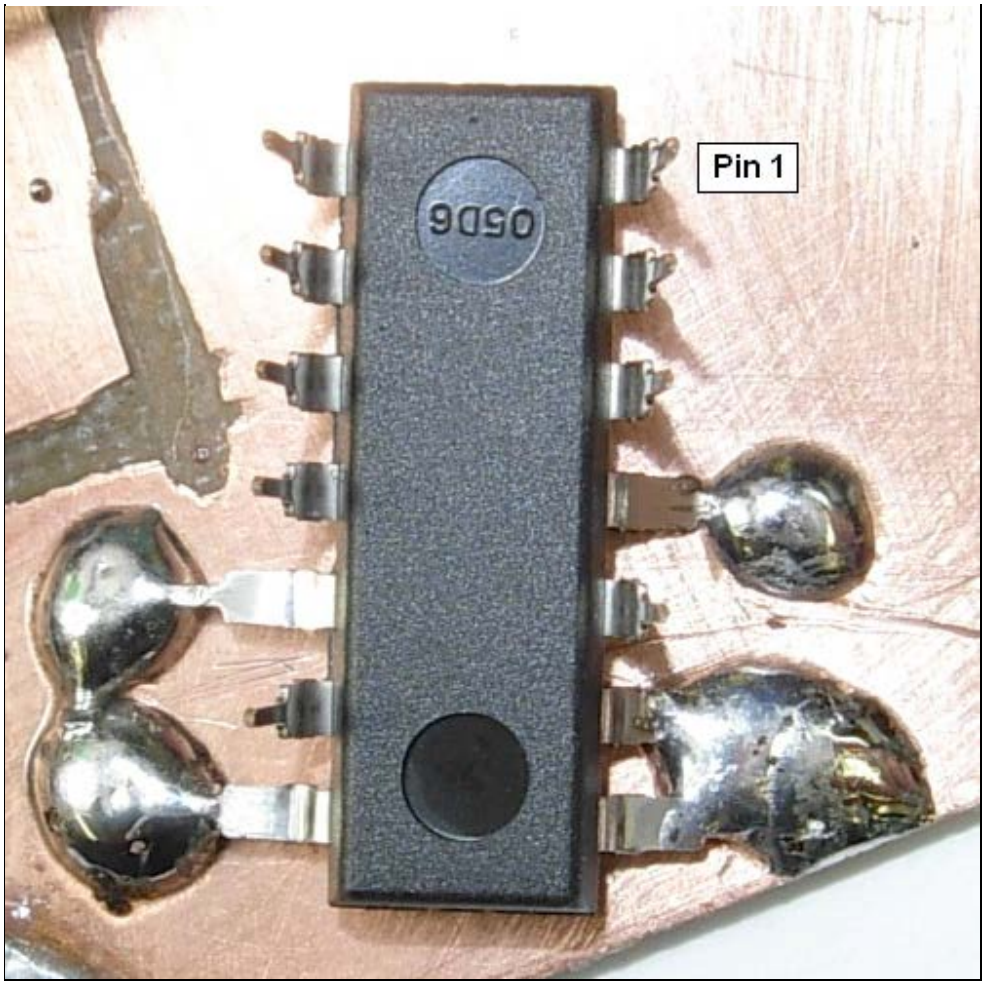
Above. The schematic of an adapted sine wave audio frequency oscillator taken with permission from EMRFD , Figure 12.4. EMRFD is the main reference for this web site. The original schematic author was Wes, W7ZOI

Above. I built this circuit from start to scope in about 25 minutes. This was a scrap, pre-used board with a positive voltage path and a potentiometer holder already on it When miniaturization is not your goal, construction is much easier and faster. I re-use parts and boards to save money. You may remove entire stages from 1 board and solder them onto another.

Above. Note how the 10K output potentiometer holder is soldered to the main copper clad board. The grounded 10K resistor is used to anchor the 22K resistors connected to IC pins 2 and 3 and can be seen in the foreground.

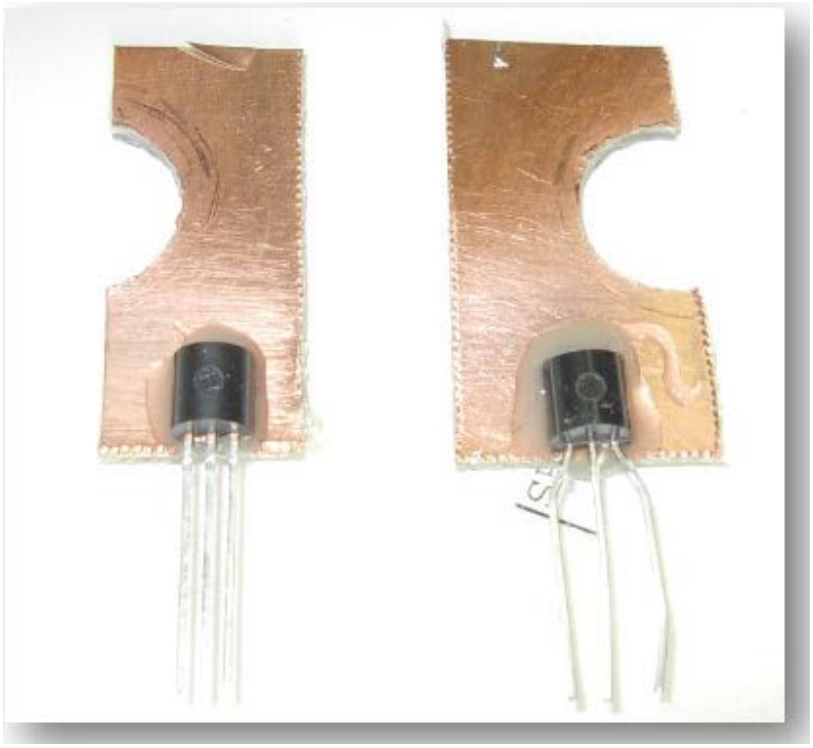


Shown above is another project. Entire control panels can be built from copper clad board for prototype circuits. In this board are numerous cut paths, 7 potentiometers, 3 jacks and a switch. Do not build an LC VFO over double sided copper clad board; lest it become "a capacitor" and affect your frequency stability.



Shown above is a CD4013B D Flip-Flop soldered "dead bug" style. Pins 4, 6,7, 8 and 10 are grounded to the copper surface; well anchoring this part. Using proper static precautions, I have never had a CMOS device failure using Ugly Construction and also save the price of an IC socket.

Above — a method to anchor op-amps using a split (negative and positive) power supply. Pins 4 and 8 are soldered to the copper board via a 10 megohm resistor. A 0.1 uF ceramic bypass capacitor is also connected to these pins. The resistor leads were left a little long to allow easy connection of the power supply wires. I write each pin's polarity on the board to avoid wiring mistakes.



Shown above is another use for copper clad board; heat sinks. In this case, 2 scraps are epoxy glued to 2N3904 and 2N3906 transistors.

A flux pen like this one from MG chemicals is a handy item for the QRP workshop. They are great for applying liquid flux to allow easy and precise soldering of SMT components. Also sometimes when adding components to ground in already built up circuit boards, it can be difficult to get your soldering iron down at a low angle for proper heat transfer. Some flux can help solder a part to the copper ground plane in these situations.

Shown above are the basic tools used to cut copper clad board. A felt pen marker, small square and a set of straight aviation shears. If you press one end of the copper board into the side of your bench and keep pressure on the handle of the aviation shears with your thigh, it is possible to make long, straight cuts. The board will flex and move out of the way as you cut. Your leg provides the force to advance the aviation shears.

The motorized grinding/cutting tool used to carve out small pathways in copper clad board.

Steel wool used to clean up copper clad board before construction. A box like this will last for years.

For 15 years, I've used this 9 mm cutter to scrape the enamel insulation off the magnet wire wound on toroidal inductors.

VE7GC Wee Willy 75 Meter DSB Transceiver Project

Introduction

I meet a lot of interesting amateur radio enthusiasts online and on the air. Among them is Dick Pattinson, VE7GC who was first licensed in 1934. When Dick isn't sailing he builds and operates his homebrew QRP gear to hams around the west coast of Canada and the U.S.

Presented is a double sideband transceiver that Dick calls "Wee Willy". This rig has a low parts count and is easily built using non-etched PC board techniques. Dick even makes his own radio case using copper clad PC board for the front and back panels and cardboard for the rest of the case. The set itself is in a case 1 1/4 by 2 1/4 by 5 1/2 inches. There is a separate container which holds a 6 volt rechargeable battery and speaker. The speaker/battery case is about 2 1/2 inches cubed and is not shown.

The circuits are built in three sections on the circuit board, namely TX, RX, and VFO. An electret condenser microphone is on front panel along with T/R switch, volume and frequency adjust. The back panel has the antenna jack, power input cord and the speaker jack. The battery pack has a 100 uF capacitor to across the power leads for additional filtering and is shown in the VFO schematic.

The text that follows is clipboard pastings from email that Dick sent me with some additional comments and expansions by me. Dick's project exemplifies practicality and innovation and with that is a major contribution for the QRPHB site.

Construction Methods

Dick's electronic construction method is quite fascinating and represents yet another derivation of ugly construction. The electronic wiring is done on single sided PC board, copper side up. Small holes are drilled through the PC board material to allow component leads to pass through them. Then the holes carrying active leads are chamfered (countersunk) with a larger drill bit which is not run all the way through the PC board. This leaves an ground-insulated side to the hole and prevents a component lead short circuit. The copper being topside allows both convenient and short component grounding. The Wee Willy parts layout is extremely neat and compact. Dick, presumably through practice has great skill with this technique and I plan to try it in the future.

The project case is constructed from 1/16th inch cardboard which is cut and bent to fit the electronic PC board. Once cut, the outer surface and edges at the front are covered with tissue paper or Kleenex (tm) type tissues soaked in white glue. The applied tissue paper and glue is allowed to dry and then additional coats are added to build up a body. Alternately, the cardboard case can be coated with lots of glue and the covering material imbedded in the glue. The air bubbles are pressed out and extra glue is added where necessary. When enough material has been added to cover up and strengthen the case joints and the glue is perfectly dry, the case is painted with Rust Coat Enamel available at hardware stores. The end result is a glossy, durable finish which looks very sharp.

Transmitter

This transmitter uses an electret condenser microphone (Dick used an Archer 270-90). The mic is built right into the front panel of the chassis and this of course guarantees short mic leads. A 741 op amp is used as a speech amplifier which in turn drives the balanced modulator a Signetics NE602 doubly balanced mixer. The input and output impedance of the NE602 mixer is around 1500 ohms.

To adjust the transmitter, set the bias control on the VN10 stage to ground and tweak L1 to resonance using an RF probe or scope on the VN10 input. The input signal must be audio, spoken into the front panel microphone to get the DSB. Once L1 is tuned, connect a 50 ohm load to the antenna with some sort of RF indicator (such as a RF power meter) and advance the bias control to give a watt or so output. Then speaking into the microphone should result in a DSB signal suitable for communicating on QRP! No audio input should result in no RF output. The supplied voltage should be kept at 6 volts, remembering that NE602's cannot stand voltage greater than 9 volts. With suitable voltage control such as a 6.8 volt zener diode on these chips, one could use higher input voltage with a corresponding RF output.

There is another way of setting the bias on the VN10. After aligning L1, with a ammeter in the six volt supply line, advance the bias control until

the input current increases about 10 mA (with no modulation). If you do not have an FT37-77 ferrite core, substitute 10 bifilar turns on a FT37-43 ferrite core for the T1 transformer

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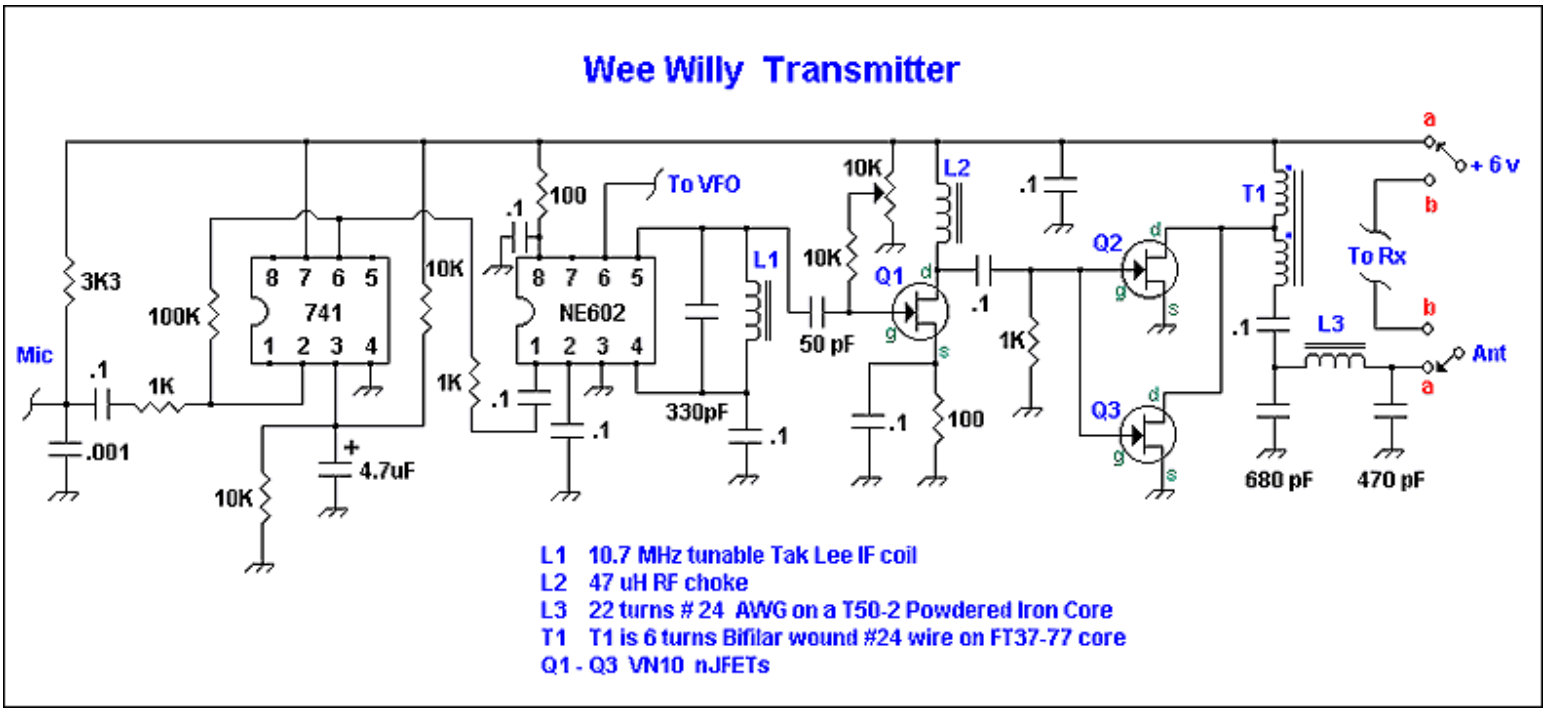
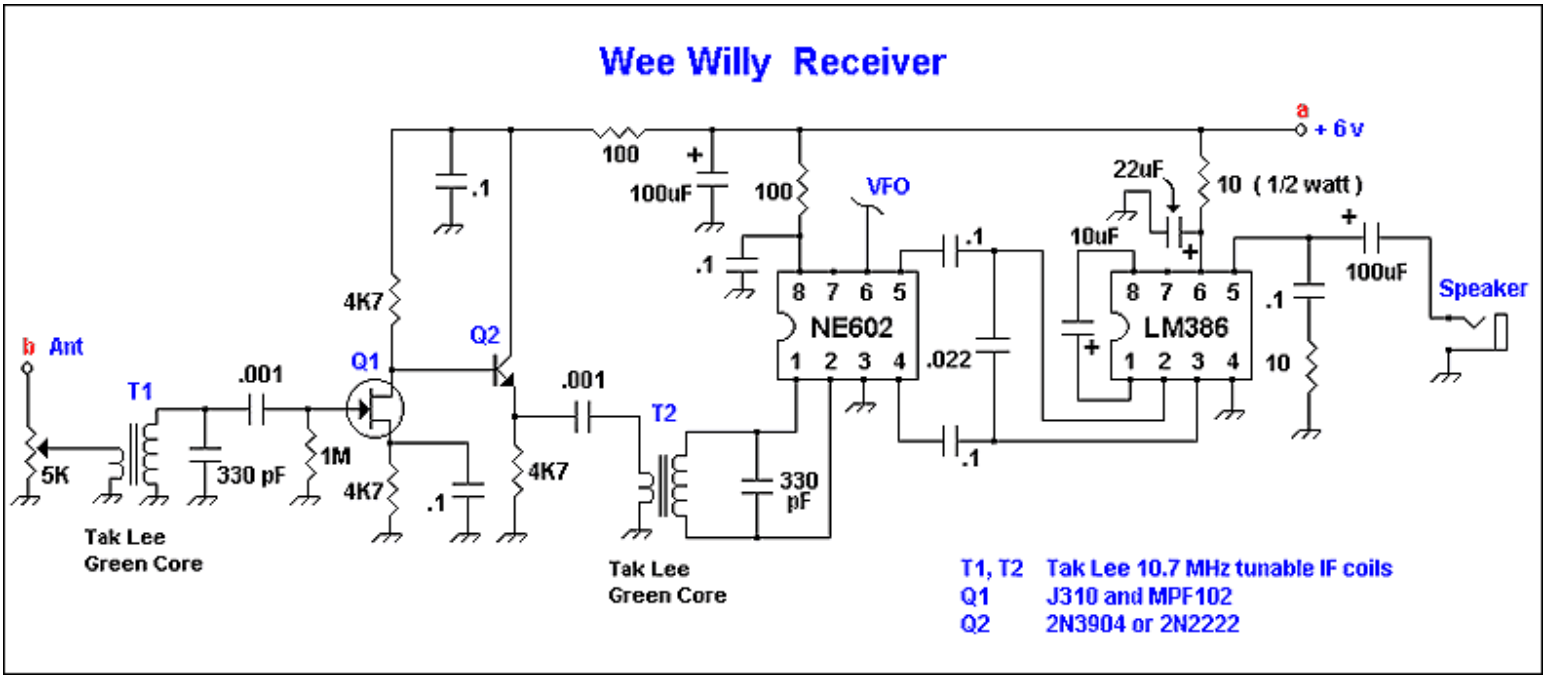


Receiver

The receiver is a direct conversion type with a manual RF gain control in the form of a 5K potentiometer. Listening to a weak signal on the desired frequency the RF stage and the mixer core (T1 and T2) adjustments are made until you hear the loudest possible signal, keeping the input test signal as low as possible. When the receiver is connected to a doublet antenna there is no lack of incoming signal, which can be controlled by the front panel RF gain control.

The antenna and 6 volt supply is switched manually from TX to RX mode and back by a front panel mounted switch. If you can not find Tak Lee green 10.7 MHz IF coils, probably any other brand of 10.7 MHz slug tuned IF transformer would work. The Mouser catalog number 421F123 would work well and in another 80 meter project I used it with a 470 pF capacitor instead of the 330 pF cap shown. I would start with Dick's 330 pf cap and if it will not tune to resonance sharply, slightly increase the cap value up to see if a bit more capacity is required to resonate it on the desired 75 Meter frequency. Note that the secondary coil on the L1 transformer in the transmitter schematic is unused. If your 10.7 MHz IF coil has a built in capacitor at the base, remove it.

During receive, the standby drain current at 6.0 volts was 24 mA and on loud signals it rose to 100 mA. If this is too much, probably the easiest thing to do would be to put in a series resistor from positive to the LM386 to limit the drain current. To get output on the speaker it is a matter of how loud you want it for the drain you draw. If earphone only reception is okay, then the drain could be reduced considerably.



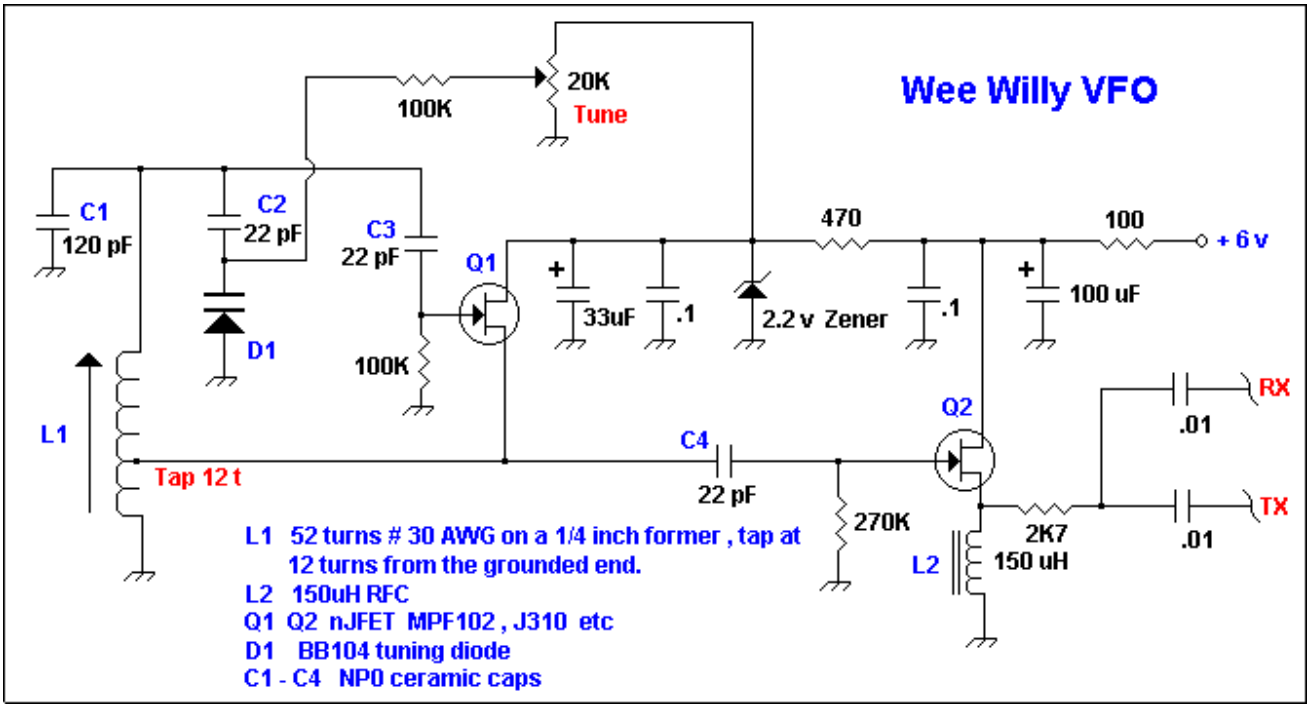
VFO

Dick's diagram indicate that this VFO was based upon a design presented in SPRAT for summer 1995. The main inductor L1 is wound with #32 AWG wire on a 1/4 inch slug-tuned coil former. This coil would have an XL somewhere between 250 - 310 ohms, so if you cannot find a coil former as described , you could easily wind one on a powdered iron toroid and make a portion of the C1 capacity variable for adjustment. A suggested alternate inductor is 53 turns of #26 AWG on a T68-6 core powdered iron core.

Dick suggests checking an old television to find suitable coil formers such as the one he used. It would probably be best to distribute the 120 pF C1 capacity among 3-4 capacitors to enhance stability. These caps should be NP0 ceramic for best results with frequency stability.

Dick's oscillator uses the slug tuned core to put VFO frequency close in frequency to where you want to operate and the variable resistor tuner on the front panel allows adjustment around the incoming signal to get the correct pitch. The desired band-edge is easily set by adjusting the slug while listening to the VFO frequency as audio on another receiver that has a frequency readout or directly with a frequency counter.

The L2 150 uH RF choke can be a simple epoxy unit which resembles a resistor. The D1 variable capacitance diode is a BB104 which has ~ 35 pF capacitance on each side. These are available at Dan's Small Parts and Kits whose URL is in the Links section of the site info web page. Experimentation with other tuning diodes could produce a practical alternative to the specified D1 part.



Operating Wee Willy

Dick sent me Willy in the mail and the first available moment I fired the little rig up and spanned the VFO which tuned from ~ 3721 to 3738 KHz. I then proceeded to tune 3729, the frequency of BCEN, our SSB provincial public service net and on my first break was able to check in with one call sign repeat to the net control station. The band conditions were noisy and most signals were S8 or lower, however Wee Willy's 1.5 watts P.E.P. were able to check me in with my folded Marconi antenna. I later changed the L1 slug and worked some stations higher up the band. My audio reports were favorible and no one knew I was running DSB. This is a fun radio and it looks cute to boot!

When transmitting, I had to be careful to keep my hand away from the VFO compartment to prevent pulling the VFO frequency with the capacitance change from my hand. The VFO has reasonable long term frequency stability and copying CW stations with the receiver was possible without frequent tuning readjustments.

The following are some digital photos of Wee Willy taken by VE7ZAC.

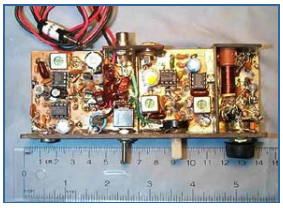
Move your mouse over the images for a larger version.

From left to right the electret mic, TX/RX switch, RF gain control and VFO tuning control. The number 375 is Dick's project identification number. He has given me many schematics of his projects and each has a unique number and case color. For the non-Canadians, the large coin on top of Willy is our "Loonie" a 1 dollar coin. The large 2 color coin in front of the rig is our "Toonie" a 2 dollar coin.

The hole on the left of this rear panel shot is for adjusting the L1 slug for the VFO. From wherever on the band the VFO is set a front panel tuning range of about 17 KHz was possible. Also shown are the speaker jack , BNC antenna connection and DC power leads.

Wee Willy with the cover off.

These top view shots clearly shows the 3 distinct compartments. From left to right the TX , RX and VFO sections. The VFO has a PC board seperator for shielding. Dick's construction method is well illustrated with this photo. The IC's are in sockets. The 3 VN10s have a small tab on the top. There are no heat sinks on the VN10 finals and they do require any for a 6 volt supply voltage. The rear panel jacks and VFO inductor are all mounted in PC board material.



Here is a bottom shot showing the connecting wiring through the chamfered PC board holes. Many thanks to Dick Pattinson, VE7GC for allowing me to present one of his projects.

Wee Wee Willy by NM8T

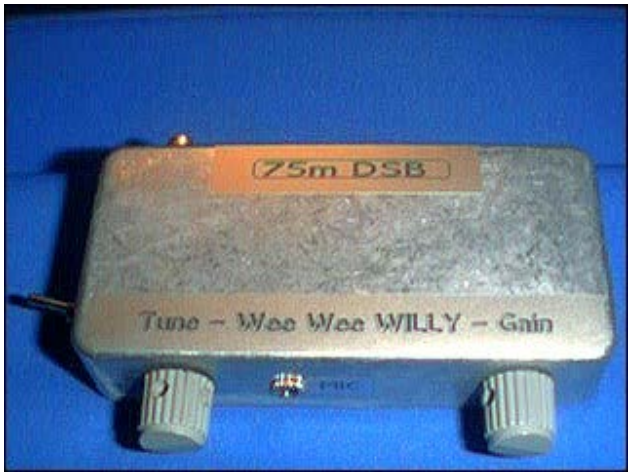
NM8T Builds Wee Wee Willy

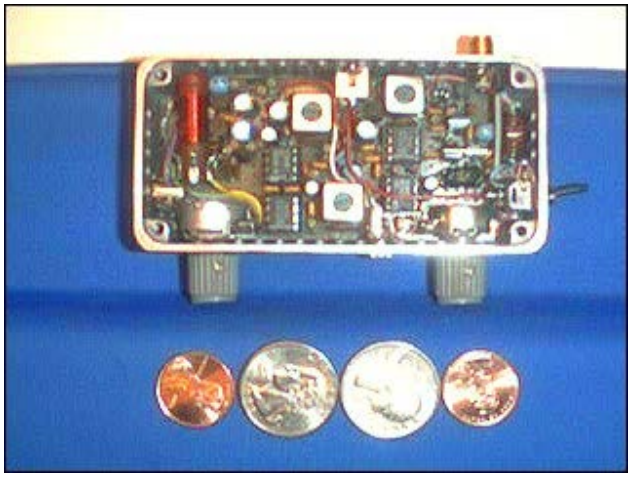
Steve White, NM8T emailed me a few weeks ago that he was building Wee Willy and was not getting the expected output power. I forwarded his email to Dick and they problem solved the issue. It turns out that the V10KM's were the culprit.

Here are their final two emails:

Steve: I think you have found a solution to your problem. I did not know that VN10's are different. I bought mine in 1994 from a radio parts supplier and they are marked VN10KM F324A1. Have fun with your new rig and let us know how you are doing with it. 73, Dick

Dick, I have completed my little rig and have had a blast with it so far. My first CQ yielded me 3 contacts in Pittsburg, Pa. My friend across town gave me a 20 over S9 report. The three hams in Pa. were astounded that I had only 1.5 watt PEP. I had to use 12 volts and used small 5 volt regulators to feed the NE602's. This works very well. I even used a little homebrew tuner with it and got my swr flat. I am going to take the little rig to Dayton Hamvention next year. I go every year and take a little project each year to show everyone. I hope it is ok to call my little rig Wee Wee Willy, since it is a little smaller than your version. I want to thank you for your help in getting this rig completed and also thank Todd for his help in getting me in contact with you. Your little rig is very neat and right up my alley for a QRP rig. I look forward to working you someday on QRP. I'm 47yrs. old on the 24th of this month. I have a whole room full of QRP rigs such as HW-7, HW-8, HW-9, Argonaut 509, 515, Powermite PM-3a. Have built a Cascade, LCK, Spider, MXM, Keyers, Tuners, and all kind of other projects. I love QRP and hope to build many more soon. The Wee Willy is one of my favorite rigs. Here are some pictures for you and Todd. Thanks and Hpe CU Agn. 73's NM8T (No more 8 Tracks) Steve White - Fayetteville, WV





My special thanks to Steve White and Dick Pattinson.

Countless builders have constructed the Wee Willy on several bands. Here are 2 more examples. Paul, KE7HR Link1 and Link 2. The finals even made it into the the BITX17 transceiver. Allan Yates published some good work on the final amp.

I've also tweaked Dick's design, but decided to leave his little rig and this web page alone out of respect to him. It's always easier to improve rather than actually *design* something. I'll never better the experiences I had operating and documenting his transceiver.

Update February 15, 2011

I lost touch with Dick and decided to call him on Feb 10, 2011. He's 95 years old and no longer building. Sadly, his wife Christina died 2 years ago. She was a beautiful person and a gifted painter. Dick remains astonished that Wee Willy - Project 375 QRP Radiophone has been built by at least 50 people (that I know about) world wide, the subject of blogs and the basis/inspiration for numerous radio projects. I met Dick on 80 meters and my wife and I once visited his residence and spent 2 glorious, days on Saltspring Island with he and his wife.

Dick sent me the radio and documentation on Aug 13, 1998. He was a hardcore builder and a professional electronics technician who, for a time, flew his float plane to service equipment in the Gulf Islands of British Columbia. Dick keeps many binders full of his carefully numbered and well documented projects. He gave me a few sheets for my reference library as a gift. Dick's work inspired me early on and I cherish the time I had Wee Willy in my radio shack - blown away by DSB, QRP and such a novel little radio.

Dick asked me to send his greetings to all QRP Homebuilders everywhere.

More Updates

Aug 25, 2011

Hi. Been viewing your web site for a while, lot of great information for builders. I thought i would let you know that I built the "Wee Willy", it works great, so good that I decided to try building one for 20m. I built the 20m version with a 14.318 crystal instead of a VFO ideal for my local net, also changed the PA circuit using IRF510 giving me 2.5W output, the transistor can run 5W, but it does get rather hot, I must say that the quality of the transmitted audio is amazing, everyone comments how good it is and some do not believe it is home-brew!!! I have put a few CQ calls out with a simple dipole and made contacts into Belgium, Netherlands, Finland and Saudi Arabia. Many thanks again for putting the Wee Willy on your website, it has opened my eyes to how well you get get out with QRP power.

Regards Wayne. M0WAY. Click for Wayne's fabulous Wee Willy bread board.

Sept 19, 2011

Wayne, M0WAY wrote again,

I built another Wee Willy and put it into a box, but with a few extra bits:

A new VFO was added that runs 14.190 – 14.350 MHz (an 11 MHz crystal oscillator + a 3.2 MHz VFO + a band-pass filter) I got the power up to 4W, however the audio was a little over powered from the extra RF, so I had to reduce the audio gain and added an antenna switching circuit, after I burnt out my receive circuit with the 4W output.

Here's Wayne's PA (from the NE602 on ** updated July 12, 2012 **) Click for a photo of his latest incarnation. Thanks again!

Oct 19, 2011

Mark, WA4JAT built a "0 cost" Wee Willy from 100% junk box parts including a CB radio transistor for the final. Click and click. Mark tunes the VFO using an air variable capacitor mounted in the former Heath Kit signal tracer project chassis. The cable running out of the front is the microphone coax with an electret mic at the end. Employing large size VFO, Rx and Tx boards facilitate easy future modifications. Awesome work Mark!

JFET BIASING TUTORIAL BY W7ZOI

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JFET Biasing and an Amplifier Design W7ZOI, 30Dec00.

Junction Field Effect Transistors (JFETs) are generally no more difficult to bias than a bipolar transistor. In many ways they behave essentially like a classic triode vacuum tube. The current flowing between FET drain and source is controlled by the gate-to-source voltage, V_{GS}. (This is the voltage you would measure with a voltmeter connected between the gate and the source.)

The Fig. 1 schematic defines the parameters we use to describe the DC behavior of a JFET. The 100 Ohm resistor is added to preserve UHF stability during measurements. The gate voltage, V, is then varied while measuring drain current. The results are then plotted, as shown in Fig. 1. The FET is characterized by the simple quadratic equation below the schematic with the two FET related parameters of I_{DSS} and V_p. I_{DSS} is the drain current when the gate is short circuited to the source. The pinchoff voltage, V_p, is V_{GS} when the drain current is essentially zero. The example in Fig. 1 has I_{DSS}=20 mA and V_p=-4 volts. Source and drain current are identical in the JFET.

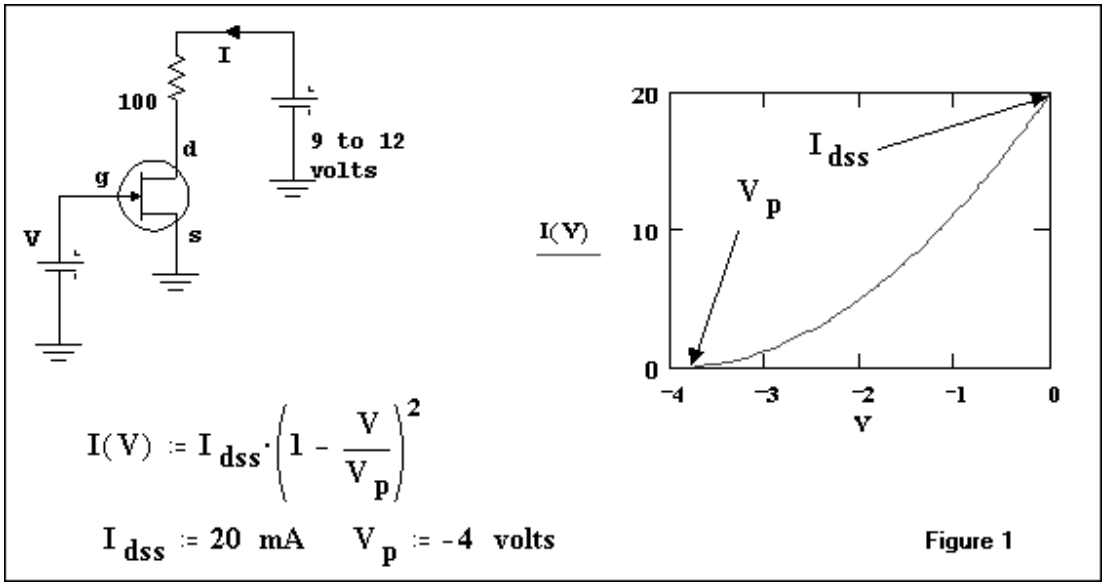


Figure 1

Basic behavior of an N-Channel depletion mode JFET.
The numbers shown basically illustrate the ideas.

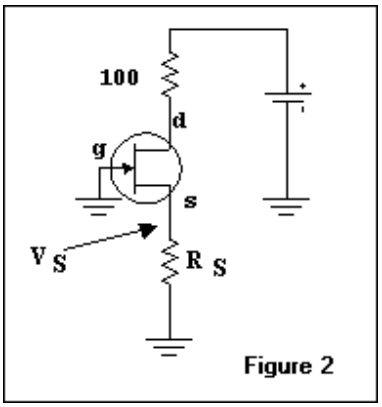
The experimenter wishing to use a JFET in a circuit will most often use the method called "self biasing" where a resistor is placed in the source lead. The gate is grounded. The FET current also flows in the resistor. The current generates a voltage drop that elevates the source voltage above that of the gate, which is the same as biasing the gate negative with respect to the source. This establishes a drain current less than I_{DSS}.

Drain voltage is not critical so long as it is a few volts greater than the source.

The resistor value for a desired bias is calculated with Equation 1 if I_{DSS} and V_P are known. But these parameters are rarely well specified. For example, the popular J310 is characterized only by ranges: I_{DSS} is 24 to 60 mA and V_P is -2 to -6.5 volts.

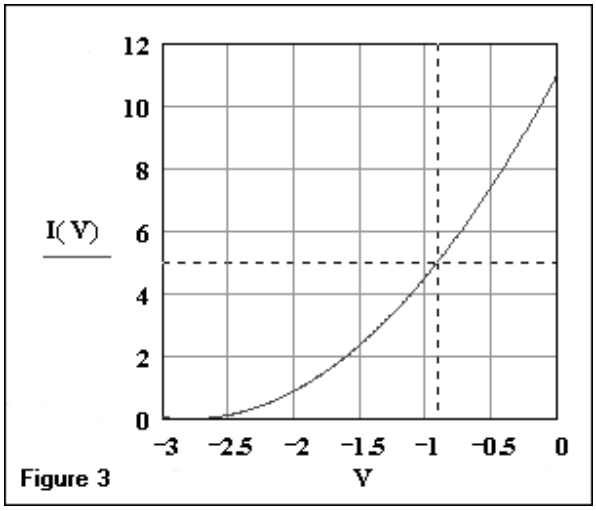
$$R_s := \frac{V_p}{I_D} \cdot \left(\sqrt{\frac{I_D}{I_{dss}}} - 1\right) \qquad \text{Eq.1}$$

I_{DSS} and V_P are easily measured for a specific FET with the circuit shown in Fig. 2. The circuit is built with a variety of source resistor values and the source voltage is then measured for each with a high impedance voltmeter. The pinchoff voltage will then be the negative of the source voltage when R_S is a large value such as 100K. Grounding the source through a current meter generates a direct reading of I_{DSS}. This brief measurement will provide enough data to allow biasing.



Circuit used to determine FET DC parameters.

This experiment was performed with a junk box 2N5454, producing the parameters I_{DSS}=11 mA and V_P=-2.8 volts. The current equation from Fig. 1 is plotted in Fig. 3 with these parameters. Additional points were measured producing points close to the curve.



Data and smooth curve for a 2N5454 we measured.
This FET has a pinch off of -2.8 volts and a IDSS of 11 mA.

This exercise SHOULD be done with virtually every FET that the experimenter uses. It is not enough to obtain data for a given FET type, although if a group of parts are available from one manufacturing batch, moderate uniformity might be expected.

We mentioned the J310 earlier with an I_{DSS} spread from 24 to 60 mA. That 2.5:1 span range is one of the better of commonly available parts. The historically popular MPF-102 has a 10:1 spread in I_{DSS} (2 to 20 mA) with V_P extending from -0.5 to -7.5 volts! Clearly, this part should be used only with extreme care. It's a good RF part, but one MPF-102 may not resemble the next.

Assume the FET we summarized in Fig. 3 was to be used in an amplifier and that we wished to bias the FET for 5 mA. The 5 mA point can be picked on the graph and matches V_{GS} of approximately -0.9 volt. 0.9 volt drop from 5 mA matches a resistor of 180 Ohms. Application of Eq. 1 produces 182 Ohms. Crosshairs indicating the 5 mA point are included in Fig. 3.

Once a FET has been biased, it can be used in a small-signal amplifier. The significant gain parameter is the transconductance, given by Eq. 2. Transconductance is the drain signal current flowing for each volt of signal applied to the gate. Note that g_M depends upon the bias current. The g_M value given in data sheets corresponds to operation at I_{dss}.

$$g_m = 2 \cdot \frac{I_{dss}}{V_p} \cdot \left(1 - \frac{V_{GS}}{V_p}\right) \qquad \text{Eq.2}$$

$I_{dss} := .011 \quad V_p := -2.8 \quad V_{GS} := -0.9$

$$g_m := 2 \cdot \frac{I_{dss}}{V_p} \cdot \left(1 - \frac{V_{GS}}{V_p}\right) \qquad g_m = -5.33163 \cdot 10^{-3}$$

Figure 4

Data used to produce transconductance for the FET used in our sample amplifier.

Evaluation of g_M from Eq. 2 for our 5 mA bias produces the data in Fig. 4.

Consider the amplifier shown in Fig. 5. We will use this as an output buffer for an oscillator. Assume that we have tapped a signal from the oscillator tank where a voltage of 1 volt peak-to-peak is available. When this is applied to the FET gate, a current of 5.33 mA pk-pk will flow. The transformer uses a 15 turn drain winding and a 3 turn output link on a FT37-43 ferrite toroid. This turns ratio transforms the 50 Ohm load to appear as 1250 Ohms in the drain line. A current of 5.3 mA pk-pk flowing here produces a voltage of 6.25 V pk-pk. The no-signal drain voltage will be about 11 volts, so positive and negative 3 volt excursions will not cause voltage limiting. The current will vary from the 5 mA DC value down to nearly 3 mA (still well above zero) and up to nearly 8 mA, still below I_{dss}. The signal at the 50 Ohm load is reduced by the transformer to 1.33 V pk-pk. This is about +6 dBm, enough power to drive a diode ring mixer. This simple circuit is a low noise, efficient, and very useful amplifier.

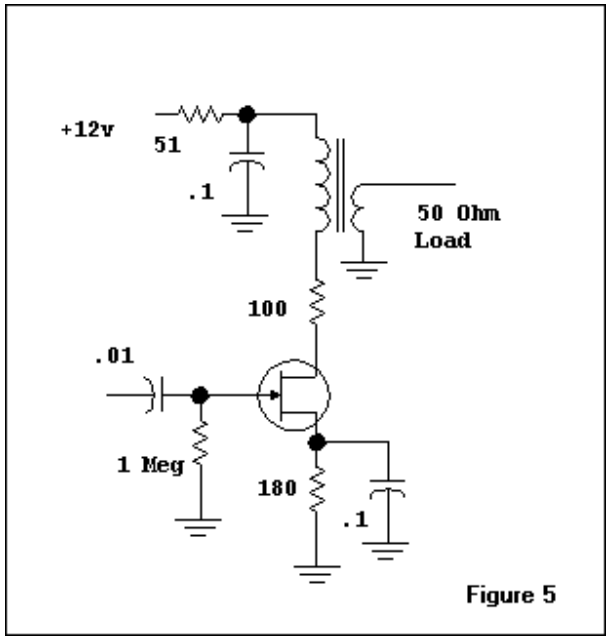


Figure 5

Common source amplifier biased for 5 mA drain current.

Figure 6 shows a FET amplifier that uses "long tail" biasing. A negative power supply is now used, forcing the biasing resistor to behave like a constant current source. This method can accommodate a wide range of FETs of a given type if the bias is set up for $I_d<I_{dss}$-min.

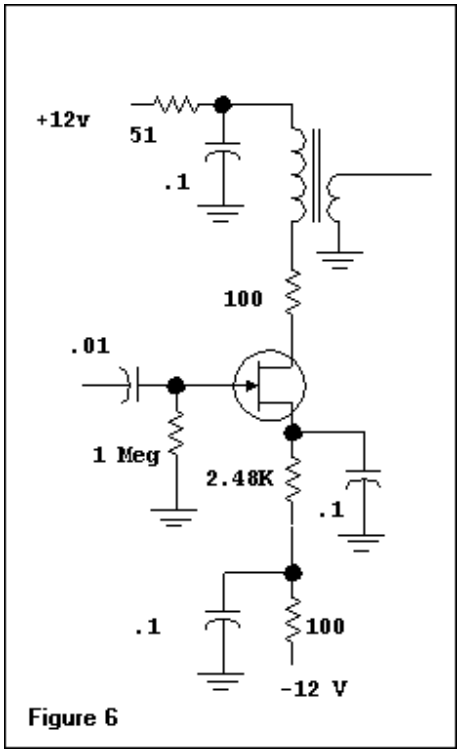


Figure 6

Amplifier with "long tail" biasing. This amplifier is biased for 5 mA and is identical in performance to that of Figure 5, but does not require the careful device characterization.

RF — Test and Measurement

EMRFD Review

Experimental Methods in RF Design

First published by the ARRL in 2003, EMRFD serves as the main QRP/SWL Homebuilder site reference.

Written by Wes Hayward, W7ZOI, Rick Campbell, KK7B and Bob Larkin, W7PUA, EMRFD is the follow-on to the 1977 ARRL publication, Solid State Design For The Radio Amateur (SSD).

A treasure trove of narrative and tabled information, schematics, references and photographs adorn this lengthy (512 page) book that comes with a data CD filled with software and key reference papers. With the included CD, EMRFD is about twice the size of the original SSD.

Wes and crew emphasize fundamental radio equipment design and bench testing rather than providing the usual catalog of circuits to just casually copy and not carefully examine.

Since we amateur builders own varied skills, abilities and test equipment, some sections may intimidate readers, while others may inspire and drive your bench practices to a higher level of competence and joy. The heart of RF design lies in measurement and reason: EMRFD emphasizes this and in doing so alienates some readers.

For some, homebrew radio electronics and kit building are synonymous — plenty of kits are sold to builders who chose to build someone else's design, rather than capture their own ideas on a scrap of copper board, or at least, to modify their equipment to suit their needs. EMRFD may appeal to builders who enjoy learning about RF *design*, or want more innovative and creative bench experiences — 1 stage at a time.

Although stressing that *build and measure* = a proven way to improve in this hobby, Wes and team share other pearls including bread boarding techniques, parts lay out, hot parts and pitfalls to avoid.

The first chapter is simply called Getting Started. On page 8, Wes shares his first simple receiver design — this page starts your breadboarding in haste! EMRFD features information and designs for all levels of experimenters, although basic electronics knowledge and some experience are needed to get the most out of its content.

Replete with sidebar examples about measuring or calculating data for common circuits, EMRFD also offers general purpose stages including a universal monoband superhet receiver front end or AF chain instead of less versatile, single- application circuits. The design information feels vibrant and flexible and some of the material is original, or presented in a way that adds to the existing amateur radio knowledge base. For example, new front-end mute circuits, or the cascode JFET mixer and RF amplifier.

Wes also shares some new ideas for RF and IF amplifiers ranging from simple to state-of-the-art. This book has something for everyone — I prefer to describe EMRFD's influence on our hobby in simple action verbs such as: improve, innovate, inspire, explore and transform.

Software:

The Microsoft Windows programs run on everything up to and including Windows 7. While a few programs are new, others represent updates of the historic W7ZOI Ladpac applications. Upon mastery of the Ladpac program suite, you'll generate the needed stages to boost *your* designs.

For example you can design complex double or triple tuned band-pass filters for your projects without any math — the software gives you the series capacitor values so you need not worry about putting links on your inductors to match input or output impedances.

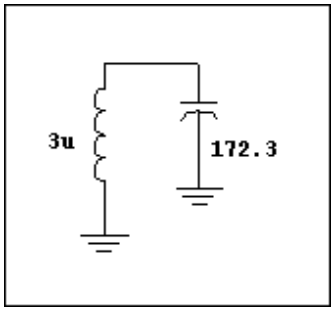
EMRFD = the **must have** reference book for your homebrew workshop.

ARRL EMRFD web link

Link to the W7ZOI errata page for EMRFD

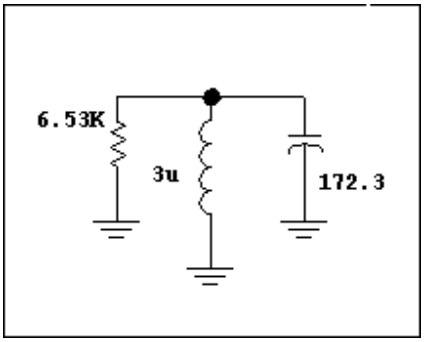
Tapped Capacitor Impedance Transformation in LC Bandpass Filters

copyright © Wes Hayward, W7ZOI, April 30, 2003.



We often use a pair of capacitors to match impedances at the termination ends of LC bandpass filters. The circuit consists of a shunt capacitor at the termination followed by a series capacitor connecting to the high Z end of a parallel tuned circuit. Some readers have asked about how the capacitors are picked. Although there is considerable flexibility in some of the choices, it is not empirical as some have guessed. A simplified double tuned circuit design sequence is presented in the sidebar on page 3.14 of EMRFD and this is the beginning of the analysis used in the programs DTC.exe and TTC.exe contained on the EMRFD CD. I'll not go into too much detail here, for it's in EMRFD, beginning on page 3.8, and in chapter 3 of IRFD. Here is what happens in the programs:

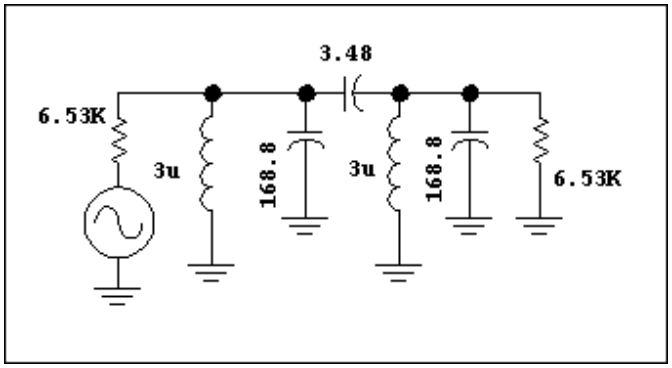
Bandpass filter design begins with an almost arbitrary choice of inductor. We pick 3 uH for a 7 MHz double tuned bandpass filter that we will use to illustrate the ideas. This L resonates at 7 MHz with 172.3 pF.



A bandwidth and center frequency are picked for the filter. This establishes a filter Q. The Q of an end section is then determined by the desired filter shape (Butterworth, etc)Let's say we want to do a 7 MHz center frequency Butterworth filter with BW=0.2 MHz. Filter Q is then 35=7/0.2. QE will then be 35x(root(2))= 49.5. The QE value in the sidebar (p3.14) includes the effect of finite inductor Q. Assume a lossless inductor for this example.

So what does this mean? It means that the end tuned circuit, when not coupled to the rest of the filter, needs to be set up to have a Q of 49.5. This is experimentally significant. (ref: QST, Dec, 1991). The reactance at 7 MHz of our 3 uH inductor is X=131.9 Ohms, so we need to load each end with a resistance R=QX=6.53K-Ohm.

What this means is that we would realize our double tuned bandpass filter with center frequency of 7 MHz and bandwidth of 200 kHz if we terminated a simple double tuned circuit in 6.5K at each end. The filter is shown below where we have used additional equations (EMRFD p 3.14) to calculate the coupling and tuning capacitors.



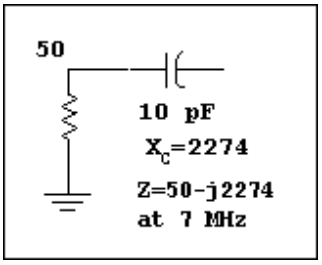
This is a useable filter design, for it will generate the desired shape and bandwidth. But, it is not very practical; it does not fit in our low impedance world.

The filter can be redesigned. One classic, but usually impractical solution is to scale the filter to lower impedance levels. For example, if we dropped L from 3 uH to 23 nH, we could directly load our filter with 50 Ohms and get the required end section Q. But this is not at all practical. First, it's difficult to build inductors with L this small and still have reasonable Q at 7 MHz. Second, the parasitic inductance of the rather large (high C) capacitors that we would attach to this inductor would begin to compare to 23 nH. A better re-design would use transformation circuits, schemes that will let us use a 50 Ohm termination (or whatever we need) and make it function as a 6.5K resistance when seen by the inductor.

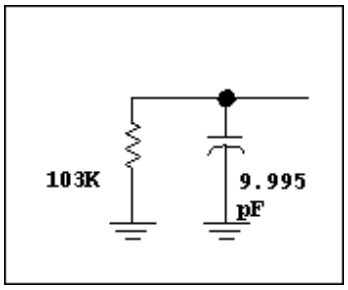
One such scheme is a transformer. This could be realized with ferrite cores or with links inductors wound on the existing 3 microHenry parts. But link coupling with design precision is a challenge of it's own. Conventional two and three element transforming networks (L, pi, Tee) are also suitable.

The simplest transforming circuit uses a series capacitor. Let's do some analysis to see how this works:

We have arbitrarily picked a 10 pF capacitor to illustrate the idea. At 7 MHz, the reactance of a 10 pF capacitor is 2274 Ohms. Hence, the complex impedance of the 50 Ohm resistance and the series capacitor is Z=50-j2274. The impedance transformation behavior of this circuit is studied by transforming the series impedance to a parallel admittance. Recall that Y=1/Z. So Y=1/(50-j2274). The result is Y=(G+jB)=9.67E-6 + j4.4E-4. Of special importance is the resistive real part of the admittance.



The 10 pF series capacitor in our example makes the 50 Ohm resistor look like a 103 K-Ohm resistor in parallel with a capacitor that is nearly 10 pF. The general case causes a R0 value resistor to look like a value of Rp with a series capacitor with reactance given by

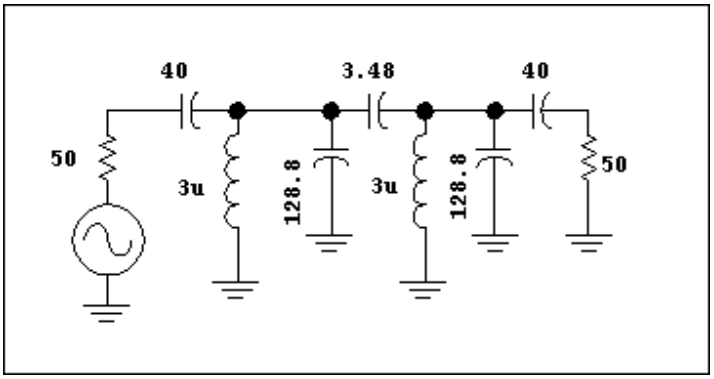


This equation is #3.1-5 from IRFD where it is derived.

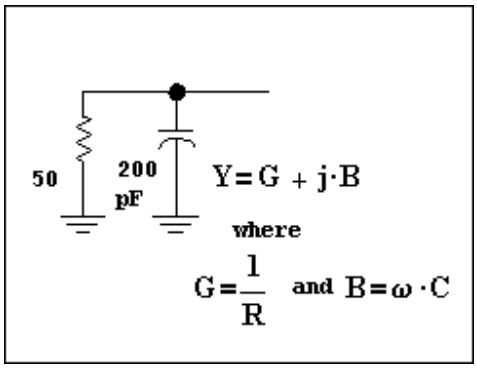
$$X_C = \sqrt{R_p \cdot R_0 - R_0^2} \qquad \text{Eq. AA}$$

Our original example filter needed a parallel resistance, Rp, of 6.5K-Ohms, which is produced by a series capacitor with a reactance of 568 Ohms. This is a capacitor of 40 pF. Another version of our double tuned bandpass filter is then

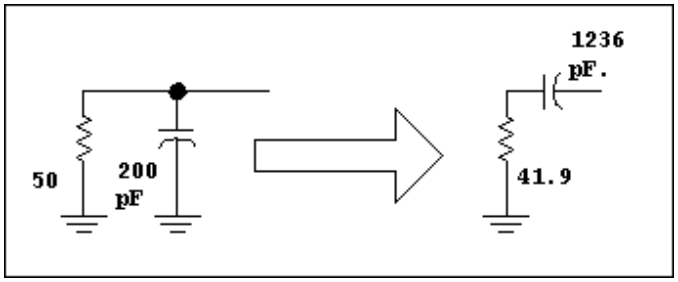
Notice that the tuning capacitors, the elements across the inductor, have dropped as C is added at the ends.

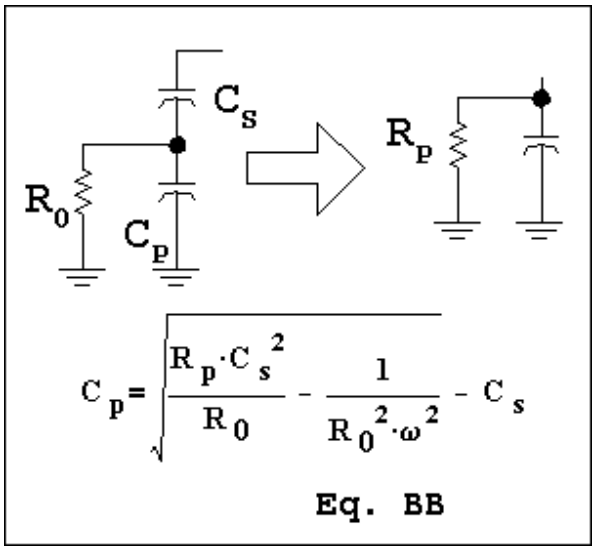


A capacitor in series with a resistance produces higher equivalent resistance. Consider a parallel combination. Here we pick a value of 200 pF as the parallel C and calculate the admittance. This is then converted to an impedance with Z=1/Y and the individual components are evaluated. This result is shown to the right.



The parallel capacitor transforms the 50 Ohms to behave like a lower value. In this case, we obtain about 42 Ohms in series with a 1236 pF capacitor.

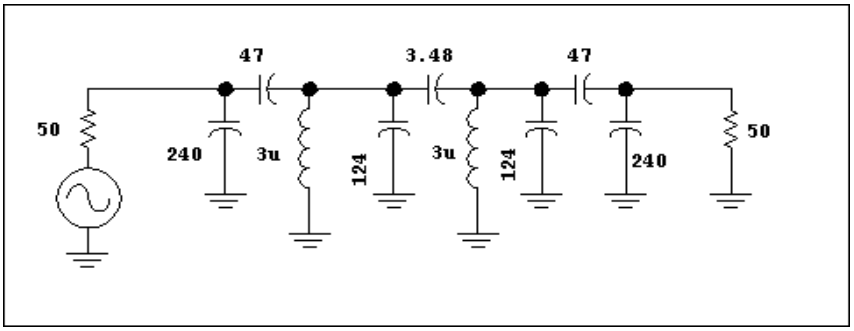




A series capacitor transforms a termination to “look like” a higher resistance while a parallel capacitor “transforms to” a lower R. Clearly, the combination of the two can generate about any result we need, realizing that they will also produce reactance that must be absorbed into the existing tuned circuits. The mathematics (now symbolic and not just number manipulation) is messy, but not difficult. A resistive termination R0 (50 Ohms or whatever) is paralleled by a capacitor Cp. The admittance is calculated and is converted to an impedance. The impedance of a series capacitor, Cs, is then added. The result is converted back to an admittance. The resistive real part is extracted and inverted to yield Rp. This expression is then solved to yield a design equation:

This is the equation used in the programs. A related expression provides the equivalent capacitance, needed to calculate the tuning capacitor for each resonator.

If we use the program with the center frequency and bandwidth presented earlier, we find that the minimum allowed series capacitance is 40 pF. If we then insert a value of 47 pF, we see that parallel 240 pF capacitors are needed to properly load the resonators. This variation is shown to the right. Circuits using the capacitor tap are practical, for they allow existing junk box parts to be used. The topology has little other utility, offering virtually the same response as a filter using only series capacitors for loading.



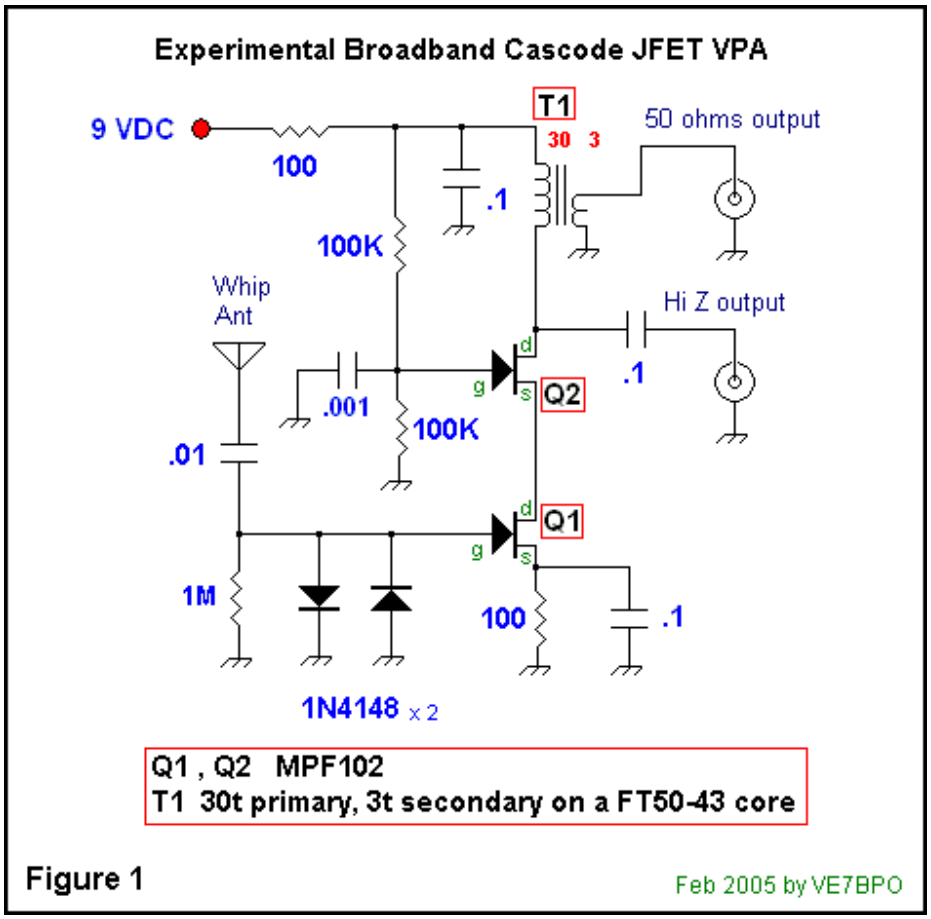
Useful link shared by Tim, KE7VYD on Feb 4, 2014. Click

More Active Antenna Experiments

Introduction

Many builders emailed me requesting a simple, broadband VPA (voltage probe antenna) design with more power gain than the common gate versions I have presented elsewhere on this web site. Connecting a whip antenna to a cascode JFET stage described by W7ZOI in Experimental Methods in RF Design is 1 method I considered.

I built the version shown in Figure 1 almost 2 years ago. This VPA. although more powerful, overloaded the front end of my test receiver with multiple RF signals. Clearly some tuning on the input was needed.



The Tuned Whip

Previous experimentation confirmed that it is easy to tune a short whip antenna by connecting it to the hot end of an L C (inductor and capacitor) tank circuit. The high impedance whip antenna was "matched" to a JFET RF amplifier by placing a high value (1 megohm or greater) on the JFET gate to ground. Although this method is practical, I desired a network to transform the output impedance of the tuned whip tank tank circuit to a known impedance. I do not possess the knowledge or mathematical skill to design such a network and asked Wes Hayward if he might consider doing this for me. My desired parameters for the network were 10.0 MHz, a 50 ohm output impedance and a 4 foot (122 cm) whip. Please refer to Wes' calculations and schematic in Figure 2 below. This math is difficult, however, a practical design for experimentation is provided.

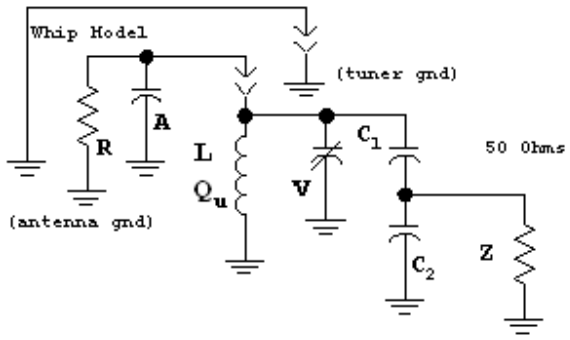
Antenna Network. The antenna is a short whip at 10 MHz. We assume a length of 4 ft with a diameter of 0.25 inch. EZnec modeled this as 7.58 pF in parallel with 3.3 MegOhm. We assumed a perfect, ideal ground for a vertical with 48 antenna segments.

$R := 3.3 \cdot 10^{6}$ $A := 7.58 \cdot 10^{-12}$ $C_1 := 47 \cdot 10^{-12}$

$C_2 := 470 \cdot 10^{-12}$ $Z := 50$

$L := 3 \cdot 10^{-6}$ $Q_u := 200$

$f := 10$ F is in MHz.



For given values of C1, C2, Z, A, R, L, Qu, we wish to calculate the value for V needed for tuning to the desired frequency (10 MHz) and the overall network Q.

$\omega := 2 \cdot 10^{6} \cdot \pi \cdot f$

"whiptune.gif"

$$R_Z := \frac{Z^2 \cdot \omega^2 \cdot C_1^2 + 2 \cdot Z^2 \cdot \omega^2 \cdot C_1 \cdot C_2 + \omega^2 \cdot C_2^2 \cdot Z^2 + 1}{\omega^2 \cdot C_1^2 \cdot Z} \qquad R_Z = 8.343 \times 10^{3}$$

$$C_Z := C_1 \cdot \frac{\left(1 + Z^2 \cdot \omega^2 \cdot C_1 \cdot C_2 + \omega^2 \cdot C_2^2 \cdot Z^2\right)}{\left(Z^2 \cdot \omega^2 \cdot C_1^2 + 2 \cdot Z^2 \cdot \omega^2 \cdot C_1 \cdot C_2 + \omega^2 \cdot C_2^2 \cdot Z^2 + 1\right)} \qquad C_Z = 4.39 \times 10^{-11}$$

The inductor resonates at f with Cnet, obtained with the usual resonance condition.

$$C_{net} := \frac{1}{\omega^2 \cdot L}$$

We must now subtract the C of the whip plus that from the load to get the variable needed to tune.

$C_V := C_{net} - C_Z - A$

Note: If CV is negative, we must reduce L or C1 to get a postive value.

$C_V = 3.295 \times 10^{-11}$

The net resistance across the inductor is the parallel combination of RZ, R(antenna), and the R related to Qu, Rp=QuωL.

$$G_{net} := \frac{1}{R_Z} + \frac{1}{R} + \frac{1}{Q_u \cdot \omega \cdot L} \qquad R_{net} := \frac{1}{G_{net}} \qquad R_{net} = 6.817 \times 10^{3}$$

The net Q is given as $Q_{net} := \frac{R_{net}}{\omega \cdot L}$ $Q_{net} = 36.167$

Figure 2

By W7ZOI Feb 27, 2005

Common Gate Amplifier Version

I built the circuit shown in Figure 3 and Figure 4 and tested it on a medium grade SWL receiver (Realistic DX 300), rather than an expensive Amateur Radio receiver. I required this active antenna for experimenting with a 10 MHz WWV superheterodyne receiver I am designing. For practical analysis, VE7TW and I did listening tests with a commercially made 4 foot telescopic whip antenna that is fitted to a standard PL-259 connector (Figure 5) and his deluxe multi-band commercial SWL antenna up a 25 foot tower.

The Figure 3 VPA was very quiet
and pulled in WWV much better

than the plain 4 foot whip of Figure 5, however, received station signal strength was quite weak when compared to the outside antenna. Our conclusion was that considering the significant losses of the 4 foot whip antenna it was connected to, the common gate RF amp does not likely have enough voltage gain to please most builders. This amp did present a low impedance to the whip network and no spurious oscillation were measured on the bench. Do not omit the 22 ohm or similar value resistor in the drain of the FET. It is used to to push the UHF parasitic oscillation tendency into the ground. Such oscillations will trash your receiver mixer intermodulation performance.

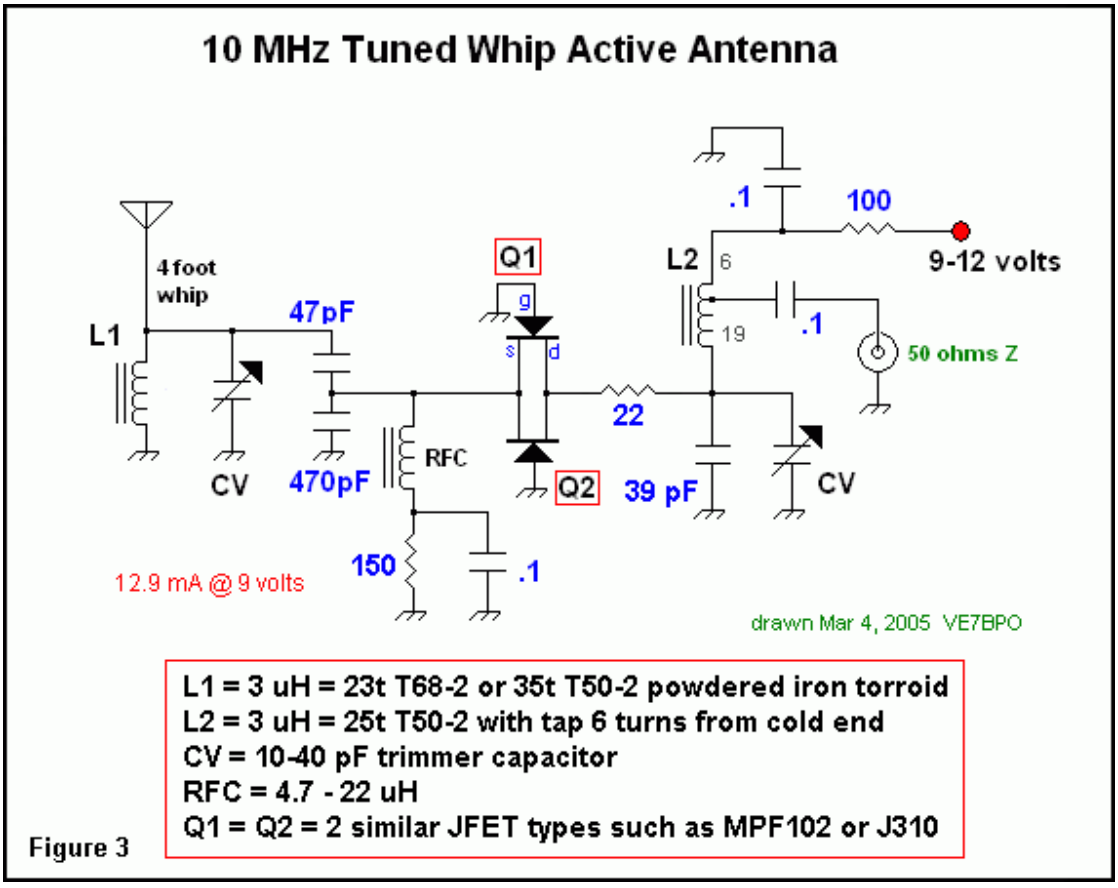


Figure 3

The 150 ohm source resistor can be increased in value and/or 1 of the JFETs removed if you wish to reduce the current draw on a 9 volt battery. It might be better to substitute 1 better JFET such as the J310 rather than use the "popcorn" MPF102 as shown. This VPA may be practical for a receiver that has an existing broadband RF preamplifier. The tap on L2 was found experimentally and the output impedance is probably higher than 50 ohms, but is likely a reasonable low impedance match to most receiver front ends. A broadband transformer for L2 might also be a good choice.

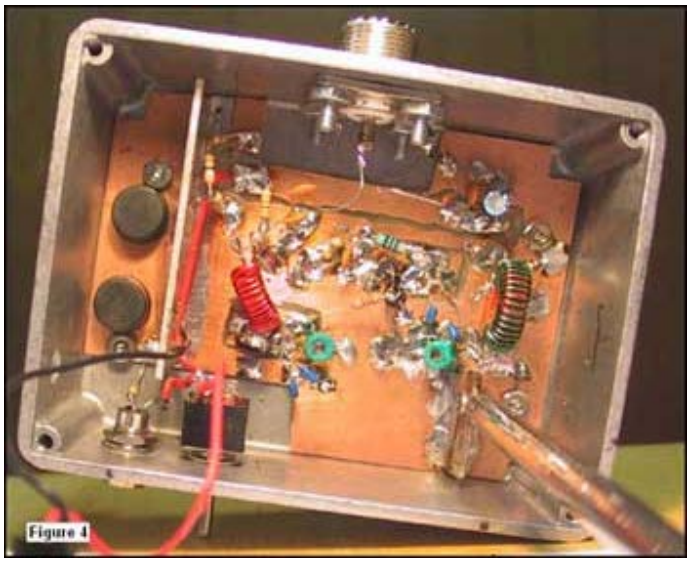


Figure 4 above: For the whip network, it is critical that you use a inductor that has an unloaded Q of 200 or above. Practically speaking, this means you cannot use a fixed value inductor such as an epoxy coated or molded RF choke. Use a powdered iron torroid instead. In my test VPA designs, for L1, I used a T68-2 core wound with (the green) 22 gauge enamel coated wire to get as high an unloaded Q as possible. If you use the T50-2 core, use 24 gauge wire if possible. Higher Q = lower losses.

Figure 5 above: This is a 4 foot whip antenna factory connected to a PL-259 that came with the Realistic DX300. It presents a very high impedance to the test receiver front end and probably wasn't a good choice to compare the VPA designs to.

Cascode JFET Amplifier Version

It was decided to use a cascode JFET amplifier to obtain more power gain. The whip network was changed to try to match the 10K input of the JFET amplifier shown in Figure 6. The whip network capacitor values (150 and 33 pF) were calculated to the best of my ability. This amplifier was tested in the same manner as the Figure 3 design. It worked very well. The WWV signal that morning was not very strong and could not even be heard with the plain 4 foot whip. The signal strength of the tuned whip was just below that of the outside antenna. The outside antenna was much quieter however and had less fading. The tuned whip antenna was quite noisy in comparison. The high gain RF amp brought up the strength of the environmental noise sources in the house. The RF gain of the receiver was reduced to compensate for the added noise.

Another problem was noted with this and other tuned drain versions of the cascode JFET amplifier; instability.

Recently, I connected a tuned drain version of the Figure 1 VPA to a receiver that contained a tuned input stage and was able to measure oscillations in the VPA with my scope. The FET drain tank and the tuned input amp seemed to be interacting.

A "swamping" resistor was placed across the VPA drain tank circuit. I had to use a resistor value of less than 1200 ohms to eliminate this instability. This greatly reduced the gain and selectivity advantage of a tuned output and I realized that output tuning may be impractical for many reasons. Some SWL builders use regenerative receivers and such a problem would be disastrous. I sent the Figure 6 schematic to Wes Hayward and he suggested using a broad or wide band amplifier as shown in Figure 8.

Instability can also occur in broadband output versions and a swamping resistor is still necessary but is used mostly to force an output impedance so that a transformer can be designed.

All of the cascode JFET amplifiers shown have fixed bias on Q2. Variable gain is possible by changing the bias voltage on Q2 with a voltage divider and/or modifying the amplifier circuit to give a greater range of bias controlled voltage gain. Please refer to EMRFD page 6.17 for information regarding this. A switchable resistor attenuator might also be practical for some builders.

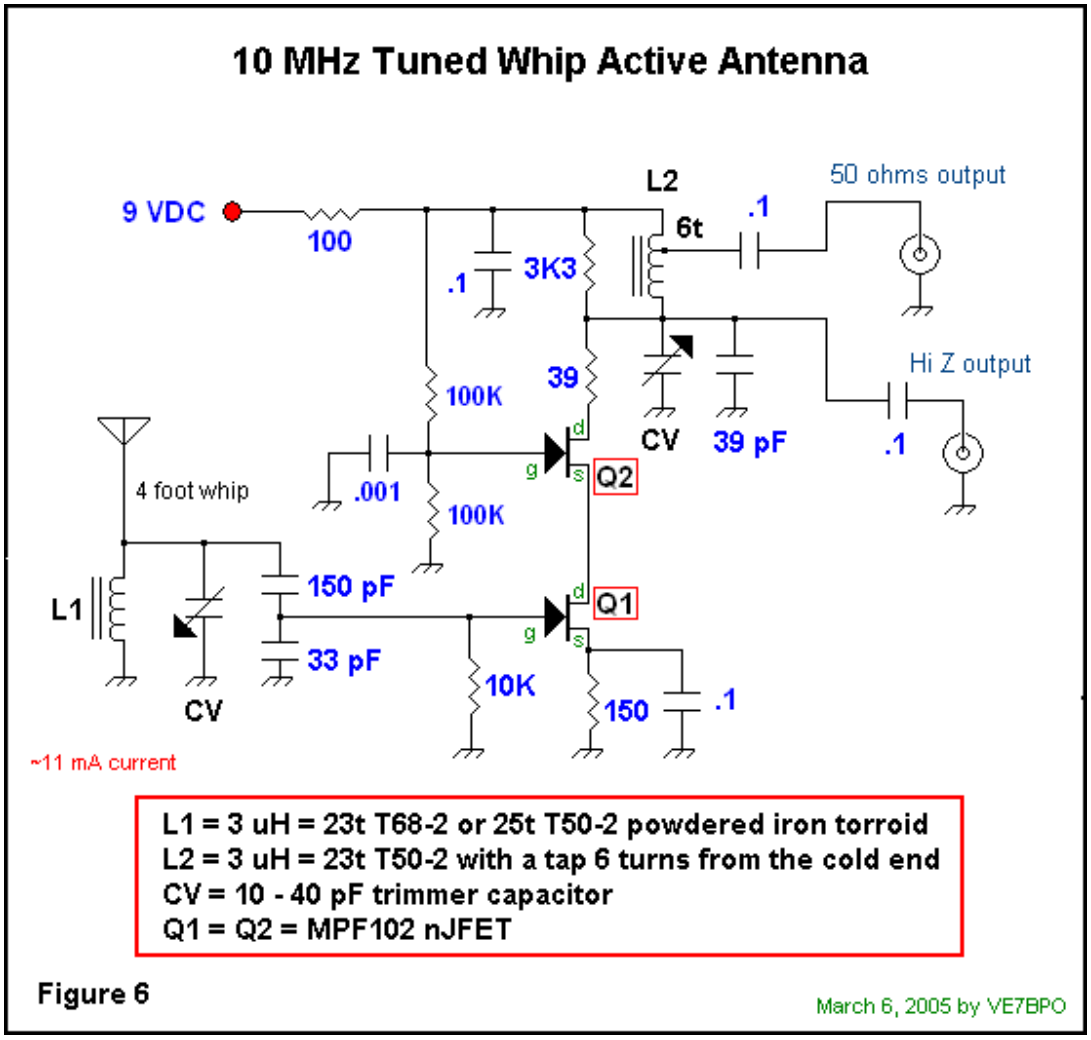
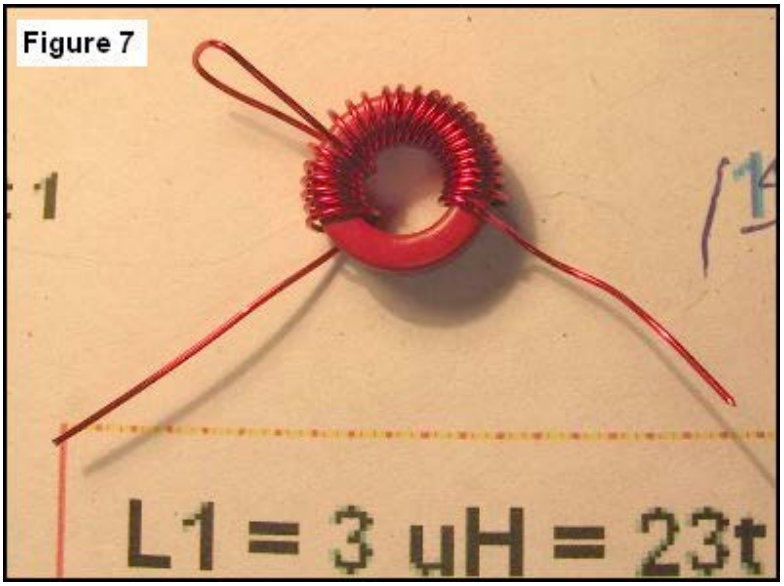


Figure 6 above: The tuned whip network is connected to a cascode JFET amplifier. A dual gate MOSFET would also be a great choice. I have many on hand, but chose the cascode JFET topology because many builders no longer have access to these devices or prefer not to use the more available surface mount types. They are also more expensive.

Figure 7 to the right: Detail of the tapped L2 inductor should you decide to experiment with a tuned drain version or need one for another project. Wind your coil and leave an extra long loop for your wire tap. Cut the tapped loop at the midpoint and use a small piece of folded ~150 grit sandpaper to remove the enamel from each of the 2 wires. Twist the now bare wires together and lightly solder them. Cut the end wires to the required length and use the sandpaper to remove the enamel. A method to strip enamel off wire is a frequently asked question for me and sandpaper works

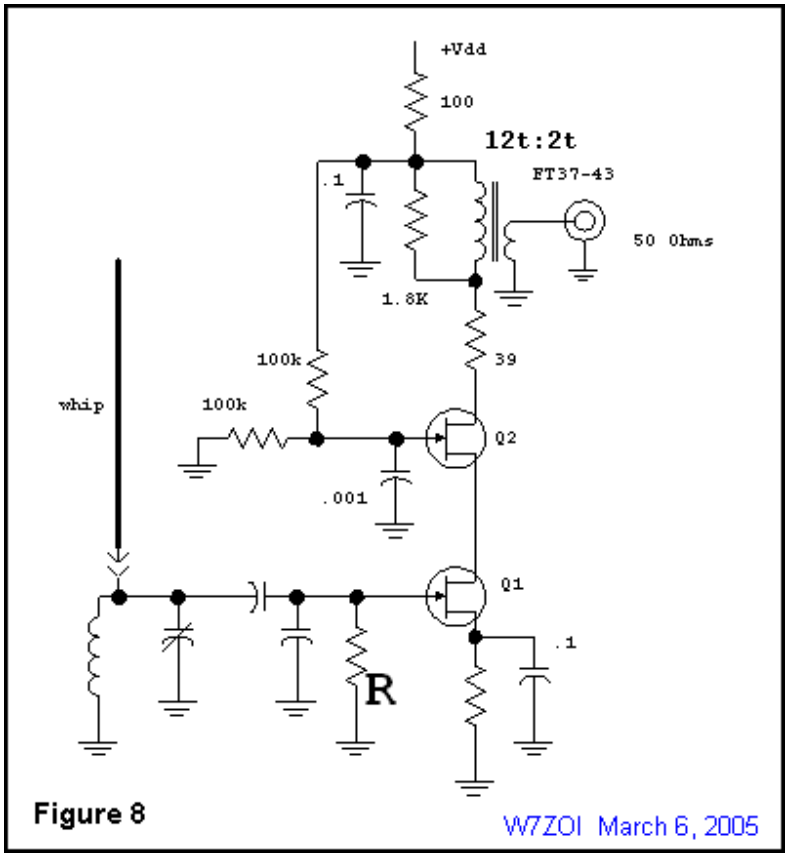


well.

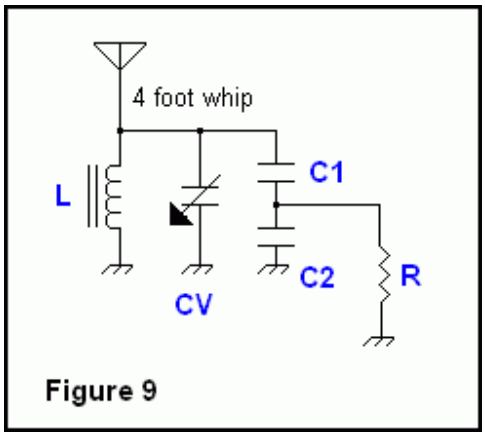
Broadband Output Version

Figure 8: The broadband version W7ZOI suggested to try building. I modified the output transformer in the Figure 6 project and tested it with Tom, VE7TW. We really liked it. By adjusting the network trimmer capacitor, I was also able to tune the 30 meter Amateur radio band as well. For 30 meter band use, I peaked the tank circuit at 10.125 MHz by listening to receiver noise with a home brew direct conversion receiver and was suitably impressed.

This is the active antenna design I wll use for my future projects where strong voltage gain is required. If your receiver has a higher impedance such as 500 ohms, you might try using a couple more links on the output transformer secondary winding.



Tuning a Whip To Other Frequencies



The ability to calculate the network values for different frequencies may prove difficult for those who lack software and/or math skills. To that end, a table follows which has some radio frequency bands and some suggested starting values for the Figure 9 parameters. Please note these are calculated and are suggested starting points based upon my limited understanding of radio electronics. Experimentation is the best method to find what component values will work for you.

Emails regarding the component values used in actual experiments is greatly welcomed.

The **R** value is the input impedance of your RF amplifier. In the case of the cascode JFET amp, it is the Q1 gate resistor.

Note that the actual circuit CV value is typically much lower than the suggested (calculated) CV value from the chart. CV is used to **resonate** the tank. CV is dependent on several factors including the capacitance of the whip antenna, your RF amplifier input capacitance , your circuit layout, component lead lengths and variations in the powdered iron core and C1 and C2 capacitors values. Expect that your whip antenna will exhibit between 8 and 15 pF of capacitance. You need to subtract this from the suggested (calculated) CV value from the Figure 10 table.

Wes, W7ZOI told me that the whip antenna capacitance will remain constant as you change frequency providing you are below 1/4 wavelength for a given frequency. Here is a good web site applet to calculate wire or whip 1/4 or 1/2 wave lengths per frequency: http://www.csgnetwork.com/antennagenericfreqlencalc.html.

The actual circuit CV might include a trimmer capacitor plus a parallel fixed value capacitor.

How to find the correct trimmer capacitor for any tuned circuit you wish to resonate

I suggest you chose a circuit CV value by placing a variable trimmer capacitor in your circuit that when set to minimum will be below half or more than the calculated CV value. Then peak the whip tank circuit using a test oscillator and scope or RF probe or by just using receiver noise. Now temporarily add a 5-10 pF capacitor in parallel with trimmer capacitor. Just barely solder it in place or even just hold it in place without touching the leads. If the output increases, you were under the correct circuit CV value. Add more capacitance and check again. Repeat until you are satisfied with your chosen circuit CV value.

If after adding the initial 5-10 pF capacitor, the output decreases, try peaking the tank again to see if you can restore the signal strength you had before you added the temporary capacitor. If after peaking, the signal strength is down, you now have too much capacitance and can remove the temporary capacitor.

You just might also have too much capacitance. You might try a smaller variable cap or reduce the value of any fixed capacitors in parallel with your trimmer to make sure your minimum capacitance is not too high to properly resonate the input tank circuit.

The point is you need to be able to tell if you have too little or too much capacitance for CV and by going under and over you can tell if you are truly resonating the tank when you adjust the trimmer capacitor. Experimentation will tell you.

Another option is to put in a front panel adjusted variable capacitor. Front panel switchable inductors might also allow other bands to be tuned with 1 tuned whip network.

Moving your body as you adjust the trimmer capacitor can change the tuning, so please keep this in mind.

Figure 10 above. Picking an inductor value for the whip network can be tricky and sometimes trial and error is required. This table may be used to find starting values for the Figure 9 network. Below 41 meters, I suggest trying a lowered RF amp input impedance as shown to allow practical component values. Most of these calculations have not been tested.

Tuned Whip Suggested Starting Values

Band	Freq (MHz)	CV (pF)	C1 (pF)	C2 (pF)	L (uH)	R (ohms)
120M	2.3-2.5	270	150	22	15	2K2
90M	3.2-3.4	165	150	10	12.6	4K7
75M	3.9-4.0	115	150	33	11	4K7
49M	4.75-5.060	83	47	22	11	4K7
41M	7.10-7.60	75	150	18	5.1	10K
31M	9.2-9.9	73	150	22	3	10K
25M	11.6-12.2	58	47	18	2.5	10K
22M	13.57-13.87	34	100	47	2	10K
19M	15.1-15.8	32	47	33	2	10K
16M	17.48-17.9	38	33	27	1.5	10K
13M	21.45-21.85	19	27	39	1.5	10K
11M	25.6-26.1	26	18	36	1	10K
10M	26.965-27.41	22	18	39	1	10K
Some Amateur QRP Frequencies						
80M	3.56	185	150	5	10	4K7
40M	7.04	85	220	10	5.1	10K
20M	14.06	56	22	10	2	10K
17M	18.069	36	33	27	1.5	10K
15M	21.06	22	27	39	1.5	10K
15M	21.06	27	18	39	.8	10K
15M	21.06	48	33	10	1.0	22K
10M	28.06	19	18	47	1	10K
10M	28.06	29	15	33	0.8	10K

Figure 10

I think an indoor active antenna for 74 meters or below might just be a noise generator.

I chose a frequency mid band for any given SWL band on the chart. The bandwidth of the input network is wide enough so this should be suitable to cover a good portion of the band.

Building An Active Antenna

To build this active antenna, chose the input tank network values from the Figure 10 chart or from your own calculations and then use them in the Figure 6 circuit. The Q1 source resistor can practically be from 100 to 390 ohms depending on how long you need your 9 volt battery to last. Increasing this resistor value will reduce the amplifier power gain. Try different values and see for yourself!

Some Practical Examples: 40 and 41 Meter Band

An active antenna that would provide coverage of the entire 40 meter Amateur Radio and 41 meter Shortwave band was designed. A varactor diode was used as the tuning element. The tuning voltage to the varactor was controlled by a 10K potentiometer which also had an integral switch. The finished VPA is shown in Figures 11 and 12.

Figure 11

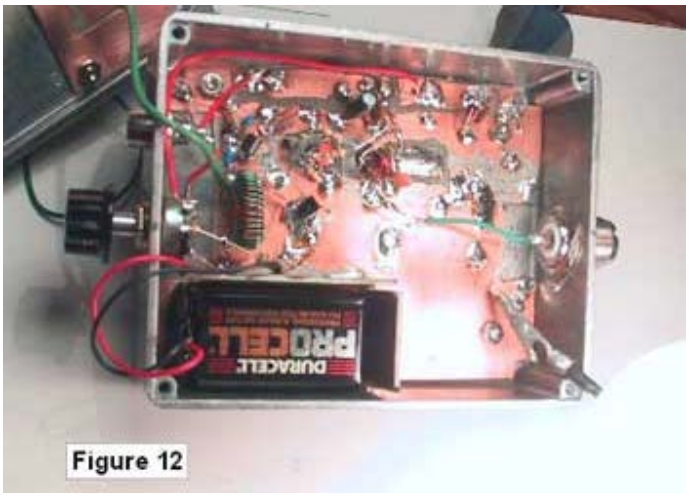


Figure 12

Figure 13 below: A hotter JFET, the J310 was used in this VPA. In the test receiver, I was able to peak a signal from ~ 6.90 to 7.60 MHz. Tuning is very sharp but peaking is easily performed by turning the potentiometer gently back and forth while listening to receiver noise or a station. It might have been better to use a smaller value zener diode as when the 9 volt battery fades below 7.5 VDC, the zener diode will not conduct and the voltage regulation will fail. Having said that, this "hotter" VPA is intended for use with an external power supply as current draw at 9 and 12 volts is 19 mA and 28.9 mA respectively.

I tried using this VPA as the antenna for the Cascode 7 receiver shown elsewhere on this web site. When the VPA was peaked at the receiver tuning frequency, loud oscillations occurred. The receiver and the VPA were about 1 meter apart. I had to turn on the -10 dB attenuator and detune the VPA for the oscillations to stop. Moving my hand near the whip antenna varied the oscillations. The high gain, tuned circuits of the Cascode 7 receiver are not a good choice for an active antenna.

Future receivers projects will have a integral VPA and clearly the front end of these receivers will have to be designed carefully. A low cost Grundig receiver was also overloaded with this VPA.

This VPA worked well with other receivers which did not have a tuned, high gain preamp.

Figure 14: From the chart, the MV209 exhibits about 44 pF (guessing) when 0 volts are

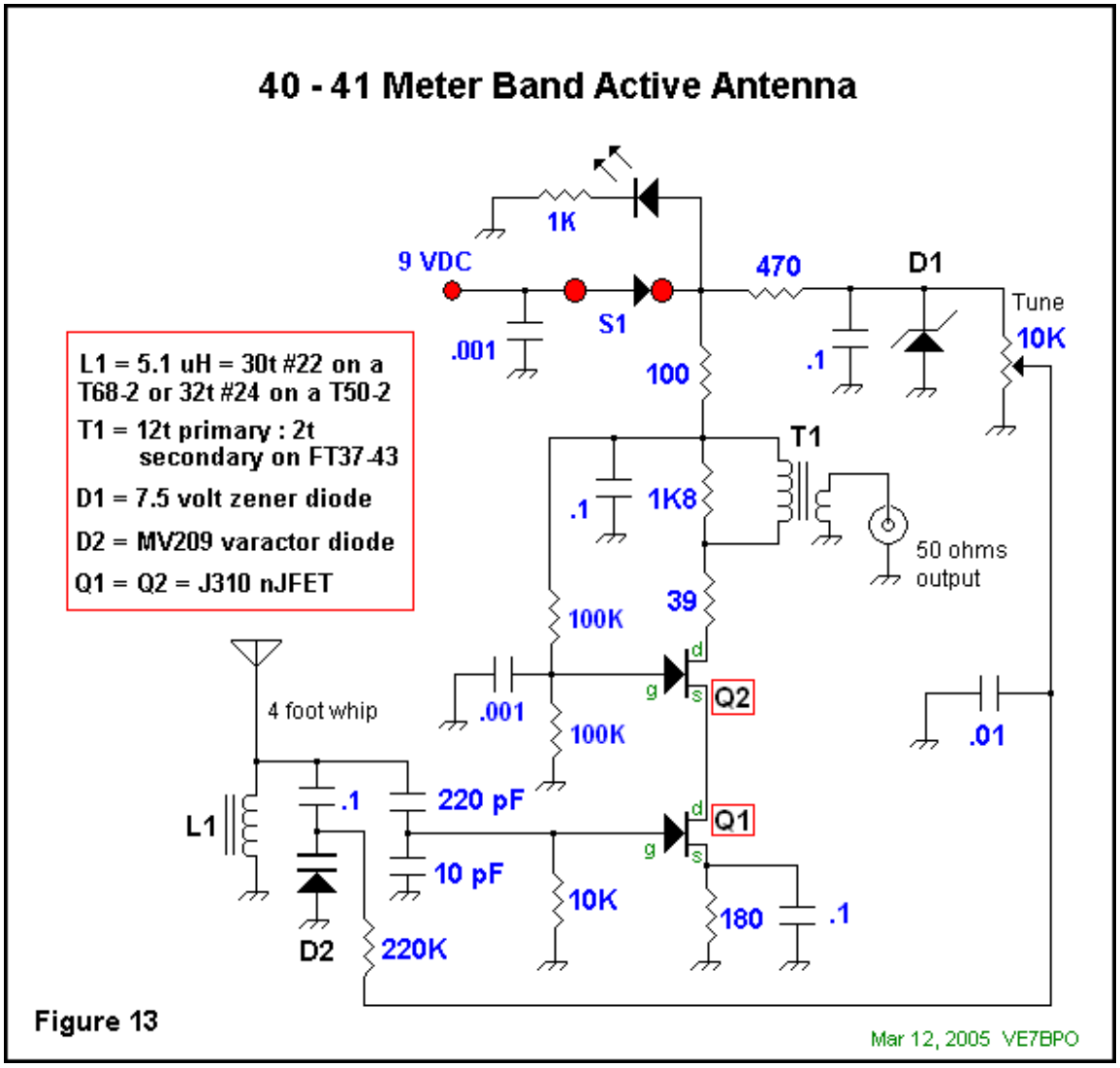


Figure 13

applied to it.

The VPA was built and tested before the tuning diode components were added. A 7.039 MHz crystal oscillator with a piece of wire for an antenna was used as an RF source. The T2 secondary was connected to ground via a 47 ohm load resistor.

A 47 pF fixed capacitor was lightly soldered in parallel with L1 and the voltage was measured with an oscilloscope. A 10 pF capacitor was then carefully held across L1 and the voltage increased by 0.25 volts. A 22 pF capacitor was then tried and the voltage decreased much below that of just the 47 pF capacitor. The nearest standard value I had on hand above 47 pF was 56 pF. The 47 pF capacitor was removed and replaced with the 56 pF one. I tried holding a 5 pF capacitor in parallel with the 56 pF and the measured voltage decreased. I had experimentally determined that to resonate L1 at 7.039 MHz I needed between 47 and 56 pF for the CV value. This range should be close enough to resonate the tank at 7.0 MHz as well.

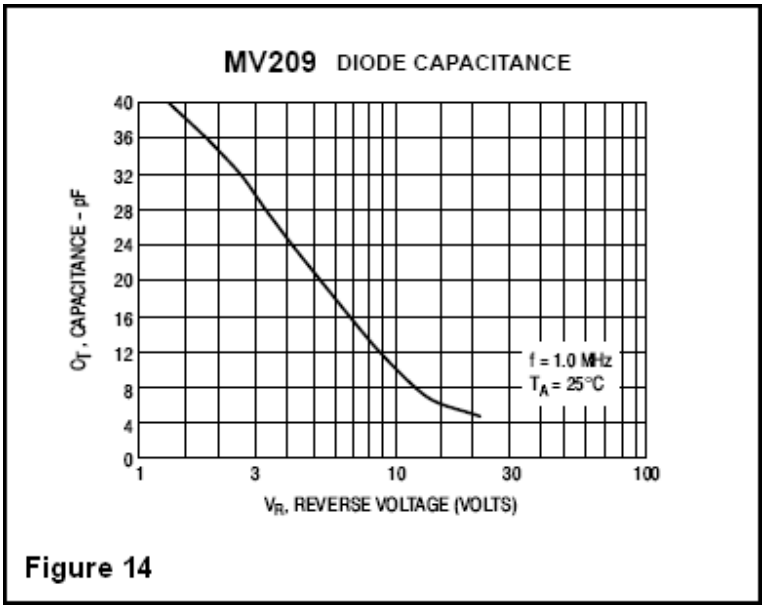


Figure 14

I then chose a varactor diode. The MV209 would be perfect for my project based upon the Figure 14 chart. I anticipated that I might have to place a small trimmer capacitor in parallel with the varactor to resonate the tank at the my lowest design frequency which was 7.0 MHz. As it turned out, in addition to the varactor capacitance, the voltage control circuit added additional capacitance and I actually needed 0.30 volts (measured between the 10K pot and the 220K resistor) to resonate the whip at 7.039 MHz. This was perfect; I did not need a trimmer capacitor! At 0 volts to the varactor diode, my whip resonated ~ 6.90 MHz.

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5 MHz WWV Cascode Bipolar Amplifier

I wanted to build a non-FET version as shown in Figure 15. The tuned whip tank was originally resonated with a 5 - 40 pF trimmer capacitor. I unsoldered this trimmer capacitor and measured it with a meter; it was 27 pF. A 27 pF fixed value capacitor was soldered in and tested. The circuit was resonant at 4.98 MHz. This was close enough for me and also the 3 high Q fixed value capacitors provided a very narrow 6 dB bandwidth along with the inductor. The output impedance value of the tuned whip is around 200 ohms to match the Q1 bipolar amp input impedance.

Listening tests indicated that this circuit probably had too much gain at 5 MHz. It might be favorable to lower the Q1 emitter current to 7 mA or so by raising the Q1 emitter resistor or decreasing the Q1 bias voltage. Also, a series feedback, degenerative resistor on the Q1 emitter might be considered. This active antenna was comparable to the outdoor reference aerial for signal strength, however, predictably was much noisier.

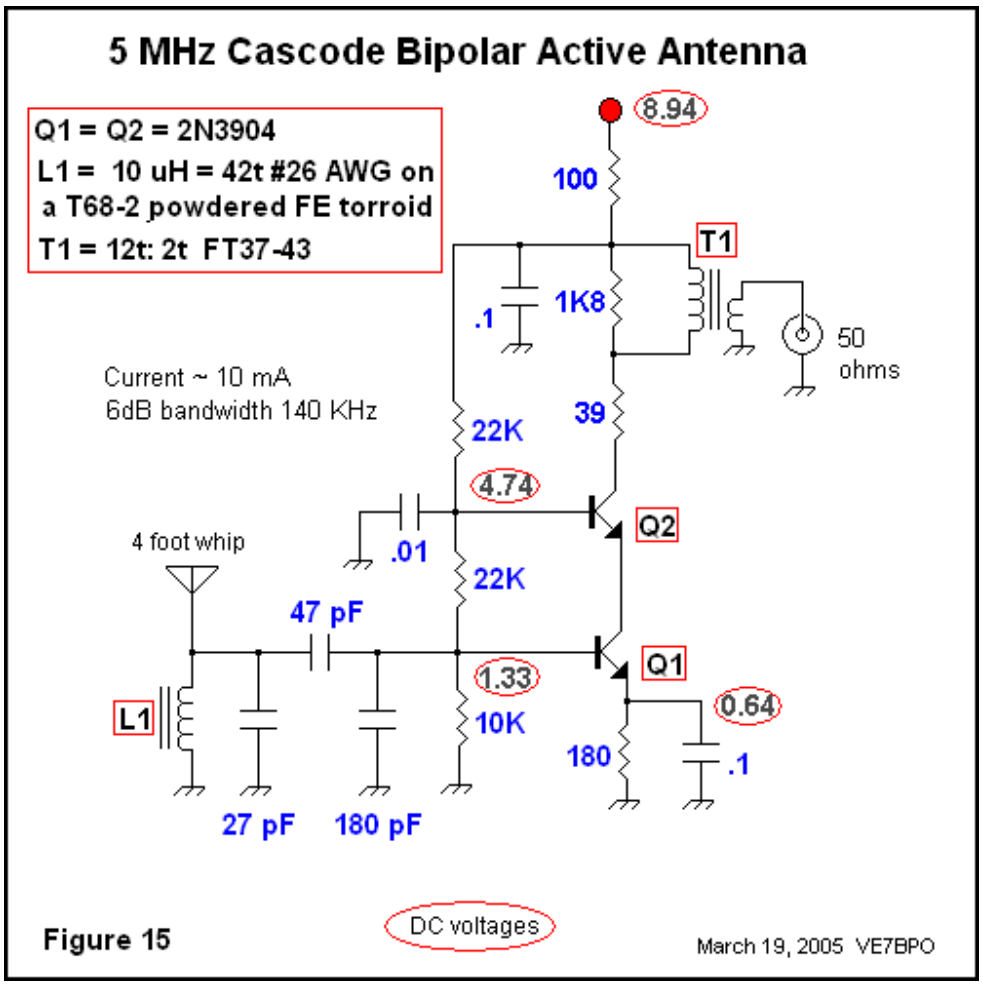


Figure 15

Final Comments

I found that using lower Q trimmer and fixed value capacitors undesirably increased the tuned whip bandwidth presumably by lowering the resonant circuit Q. The inductor unloaded Q was the dominant factor however. The worst case scenario was a tuned whip built with junk-quality parts which had a -6 dB bandwidth of ~390 KHz.

I also learned that you should expect high gain amplifiers to oscillate and specifically design to reduce or suppress this tendency.

The 2005 Active Antenna experiments were fun and provided many learning opportunities. An active antenna is a perfect weekend project. There is no substitute to learning by building and testing electronic circuits with your own hands.

My sincere thanks to all of the friends who helped me with these experiments.

Experiments by Other Builders

What follows are some VPA experiments by others that were sent to me by email. I seek your feedback and photographs to help improve this web site and also to gain motivation to add more new content.

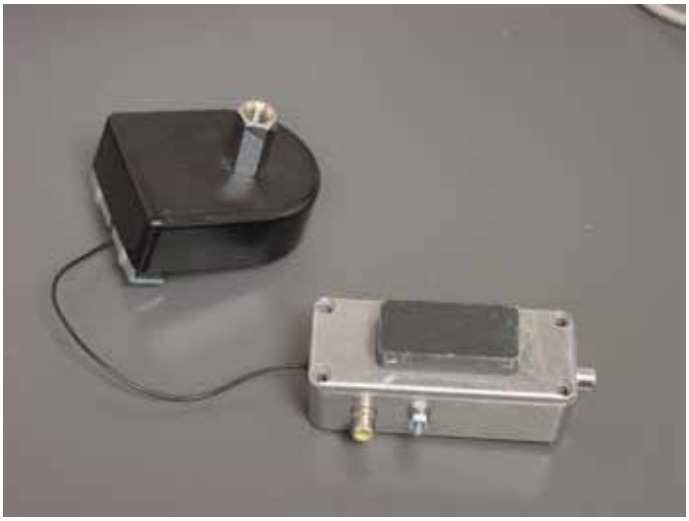
Joe, K9LY
Hi, Todd :

Attached are some photos of a voltage probe amplifier that I built using ideas from your excellent website. I use a TenTec 1254 receiver in the car and listen to some shortwave broadcasts during the daily commute to and from work. The antenna is a 4-foot whip that screws into the trunk-lip mount shown. The amplifier is held to the bottom of the trunk lid by a magnet and has survived for several weeks without falling off.

The amplifier is tuned by a varicap diode and covers approx. 9-14 MHz. The tuning voltage comes from a potentiometer that I added to the front panel of the receiver.

I decided to use the 2N3904 cascode amplifier because I liked the idea of using the most common transistor possible. The LC-tuned input is nice because the antenna whip is held at chassis ground potential, which should help prevent damage to the amp caused by static buildup.

The TenTec 1254 Receiver. http://radio.tentec.com/kits/Receiver



Above left: Inside Joe's trunk lid SWL Active antenna. Great ugly construction in a sturdy Hammond chassis.
Above right: Joe's remotely tuned SWL active (or voltage probe) antenna amp and whip holder. Thanks Joe!

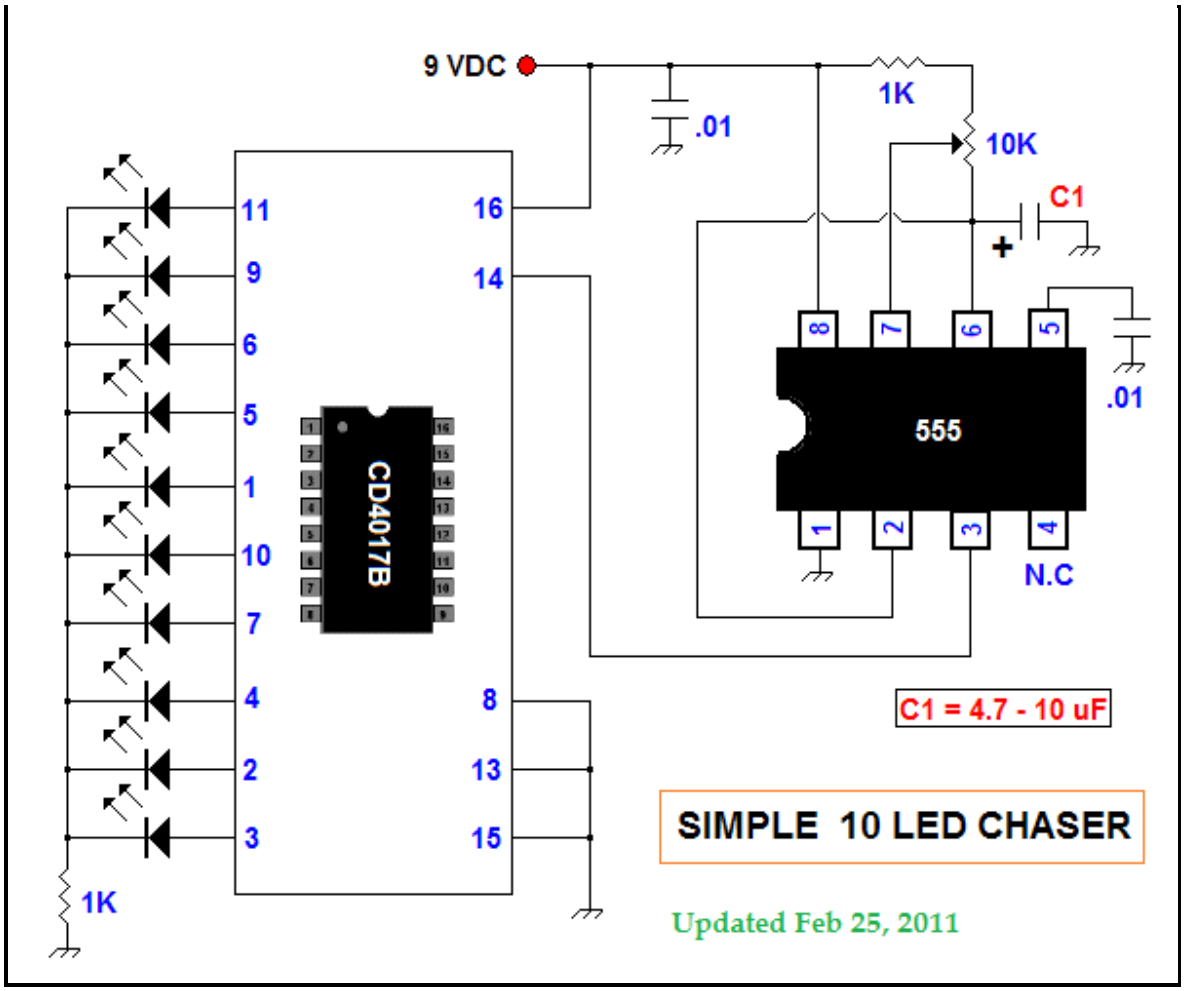
Fun with LEDs

Introduction

This summer I built several LED projects including sequential LED chasers (sequentially left to right) and also "Nite-Rider" style which go (sequentially left-right-left-right-etc). Many more LED schematics can be found on the World Wide Web via a Google search. LED projects are great fun for both HAM's and SWL's alike. They are also a lot of fun for children to experiment with. Currently, I am experimenting with PIC microcontrollers to perform LED "tricks".

I also built several very bright LED flashlights which run on a single 1.5 volt battery. For ultra-bright LED flashlight schematics, check out Dick Cappel's excellent and very informative web site. He has a number of LED driver circuits and other great schematics and theory. To wind the inductor for these LED flashlights, I had good success using an FT-37-43 ferrite torroid core. I used at least 40 turns of wire which is generally center tapped.

LED Chasers



Above. This is the schematic for a ***very*** basic 10 LED chaser I built. I prefer my "chasers" to run slower than most and chose a 10 uF capacitor for C1. The 10K pot can reduce the flash speed from not moving to whatever minimum time constant is possible with the C1 value you choose. Don't bother with ultra-bright LEDs for these "chaser" projects as cheaper, lower millicandela (mcd) LEDS work fine. I favor blue and green LEDS. The 4017 decade counter is a fabulous part and can be driven to flash a row of LEDs with a 555 timer chip or a discrete BJT multivibrator.

Update Feb 25, 2011: Many "experts" have emailed, or flamed me on web forums to say this circuit can't possibly work. I really hope these unhappy men cultivate enough humility to study and understand the 555 and more importantly; to reap some of the happiness and joy that comes from being positive and helpful to others. Since 2005, greater than 300 builders have emailed to say this simple circuit works and they want to learn more about electronics.

My intent was to have the least number of parts to flash some LEDs. Some new builders become overwhelmed when the parts count is high — I once shared this fear and relate. You'll see a number of different bias circuits for the 555. Many builders run the reset pin; **Pin 4** high (connected to the 9 volt battery) and as a rule, this is a good thing to do, but it's not necessary for the circuit to work. Pin 7 is an open collector output to ground — its primary purpose is to discharge the capacitor.

It's important for the DC voltage in the pot wiper to not become too close to the + 9 volt rail or VCC (This happens when when the pot is rotated so that maximal DC voltage appears on pin 7), as pin 7 would draw excessive current. In my original schematic I left out a series resistor from VCC to the pot to eliminate this problem. After some thought, I added a 1K resistor on Feb 25, 2011, although this limits the rate somewhat. I run the pot on my circuit about mid-range and it hasn't been re-adjusted (or turned off) since 2005.

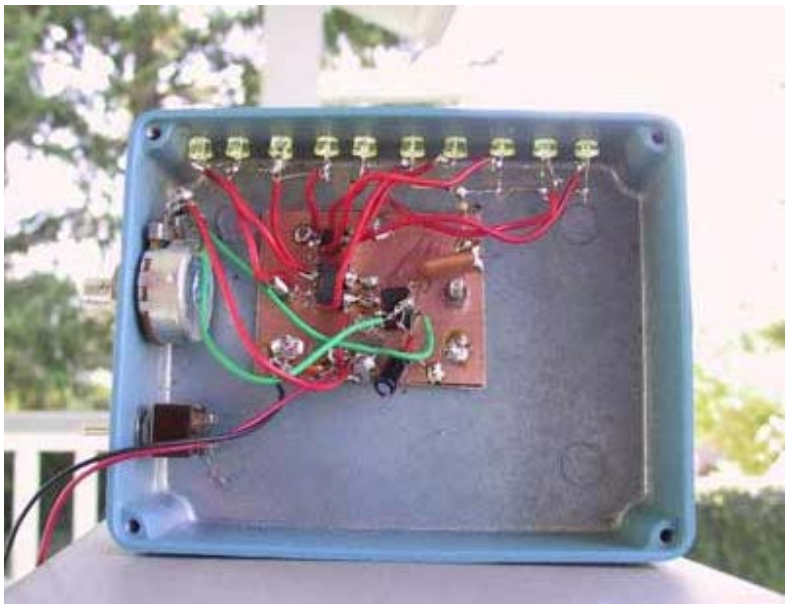
Also, the rotated pot wiper shouldn't get within a couple of hundred ohms of the capacitor as that too would cause excessive current spikes into pin 7. Generally, I prefer not to have much current on the wiper of a pot or, at least, try to keep the current small. Often, you can use a pot to set the desired timer speed and then remove, measure and substitute 2 standard value fixed resistors. A better way is to use math and calculate the resistor values, but this involves math and some people want nothing to do with equations.

This circuit is meant to provide a minimalist working circuit, but doesn't provide a great example of 555 design. Happily, for those wanting to learn more, countless great 555 tutorials may be found on the web. One of my 555 favorite sites is that managed by fellow Canadian Rob Paisley. Increasingly, I am exchanging electronics-related emails with model railroad enthusiasts across the globe and many of them know of Rob's wonderful web site.

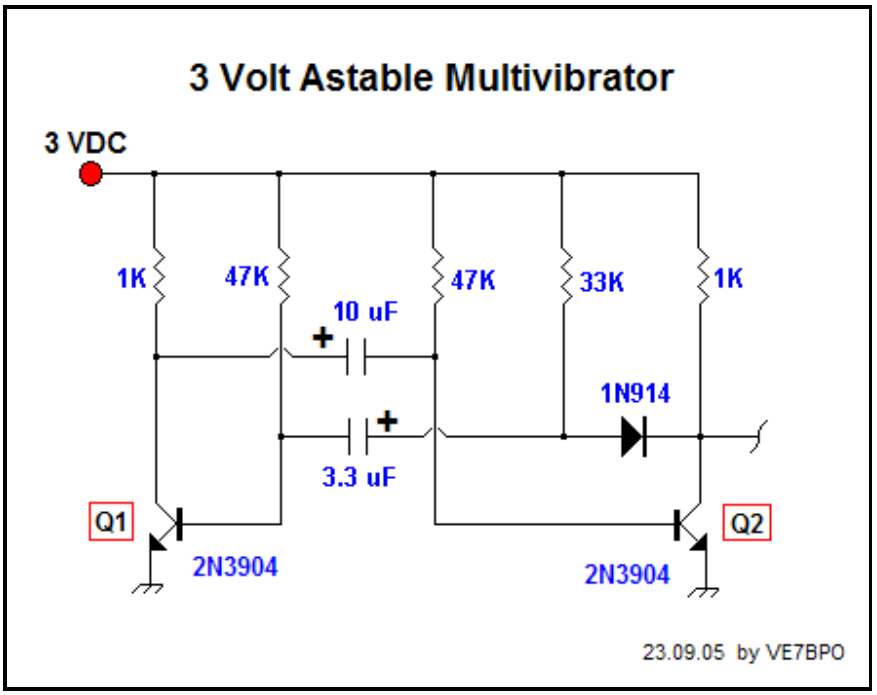
Matthew Ritchie built and posted a nice version of the LED Chaser on YouTube. A reader sent in this breadboard photo.

By far, the coolest device incorporating the LED chaser lies within a sculpture called Cyanic by Seattle-area artist Allet. Click for his web site. Cyanic may be found on the New Sculpture Build section on his web site. You have to start the Quick Time video manually with a mouse click. I love Allett's work and his lastest light sculpture exemplifies how the Internet can unite creative people with a positive attitude.

A 10 LED sequential flasher in a blue Hammond chassis. The schematic is shown above.

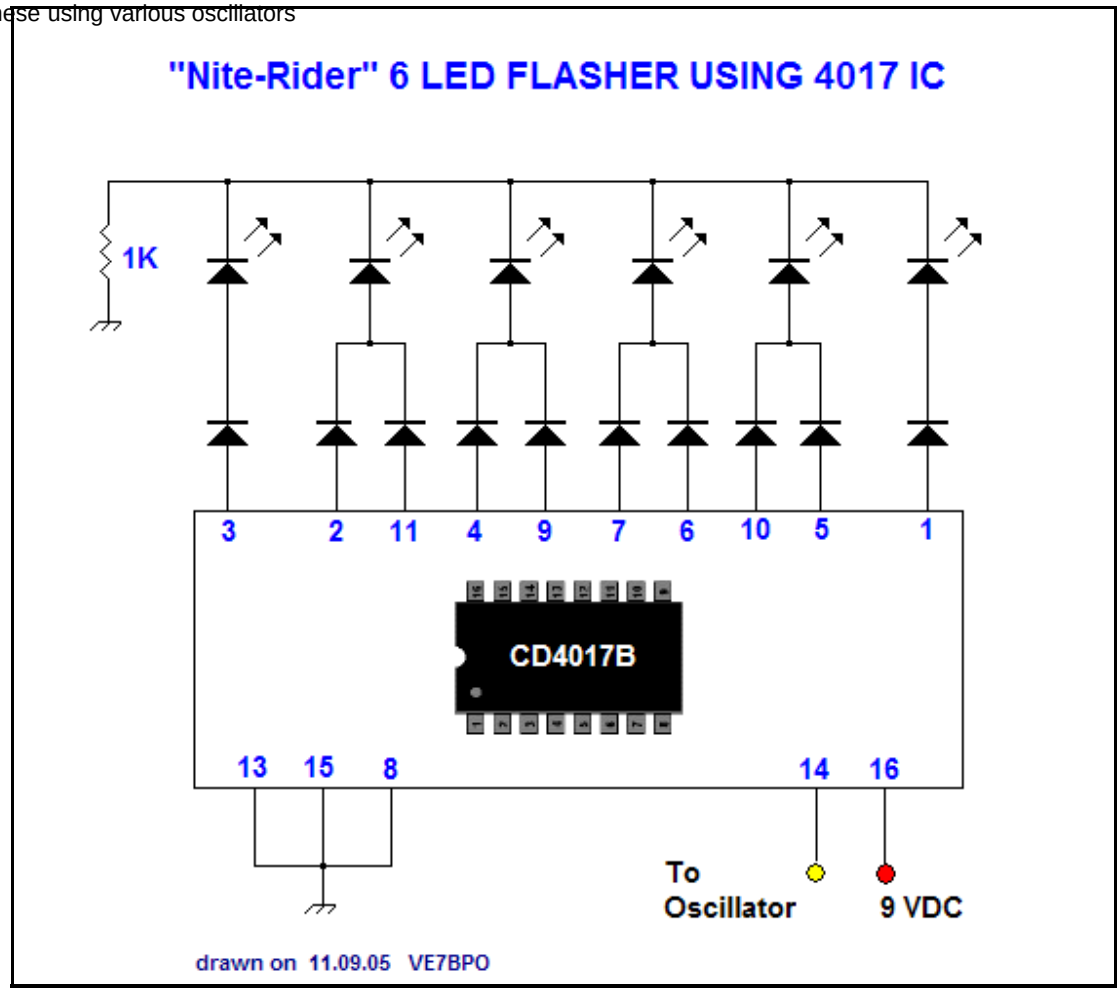


An RC oscillator designed for a 3 volt LED chaser. It oscillates quite slowly so the LED chaser it triggers will not be overly distracting. Some RC oscillator design details are discussed later. This oscillator triggered a 4017 decade counter instead of the 555 timer chip shown in the "Simple 10 LED Chaser" schematic. There are many links describing the theory of the 2 transistor astable multivibrator on the World Wide Web. I also have some information on this web page.



Above . This is a tiny 3 volt chaser which uses an LED bar instead of discrete LEDs. It draws 3.8 mA peak current on pulses. It uses the optimized BJT astable multivibrator shown directly above which fires at ~120 cycles per minute (slowly). The 3 volt battery pack is hidden behind it and should last several months. Soldering the LED bar was not an easy task. The plastic Hammond case measures 2.46 by 1.38 inches (6.25 by 3.5 cm).

A schematic to allow the 4017 decade counter to sequentially flash 6 LEDS left-right-left-right-etc. Connect your favorite square wave oscillator to pin 14. I built 4 of these using various oscillators and LED colors. You might consider using lower DC voltages and if so, may adjust the 1K current limiting resistor by using ohm's law. The 10 small signal diodes may be any appropriate type including the 1N914 or 1N4148. None of my 4 projects exceeded 6 mA peak current draw, so battery life is excellent. I increased the 1K resistor to 1K5 in my 4th project as I found the LED's that I used too bright.



The prototype "nite-rider" project with messy wiring. The holes for the LEDs were bored with a hand drill and it shows! The discrete transistor multivibrator can be seen behind the 4017 IC.

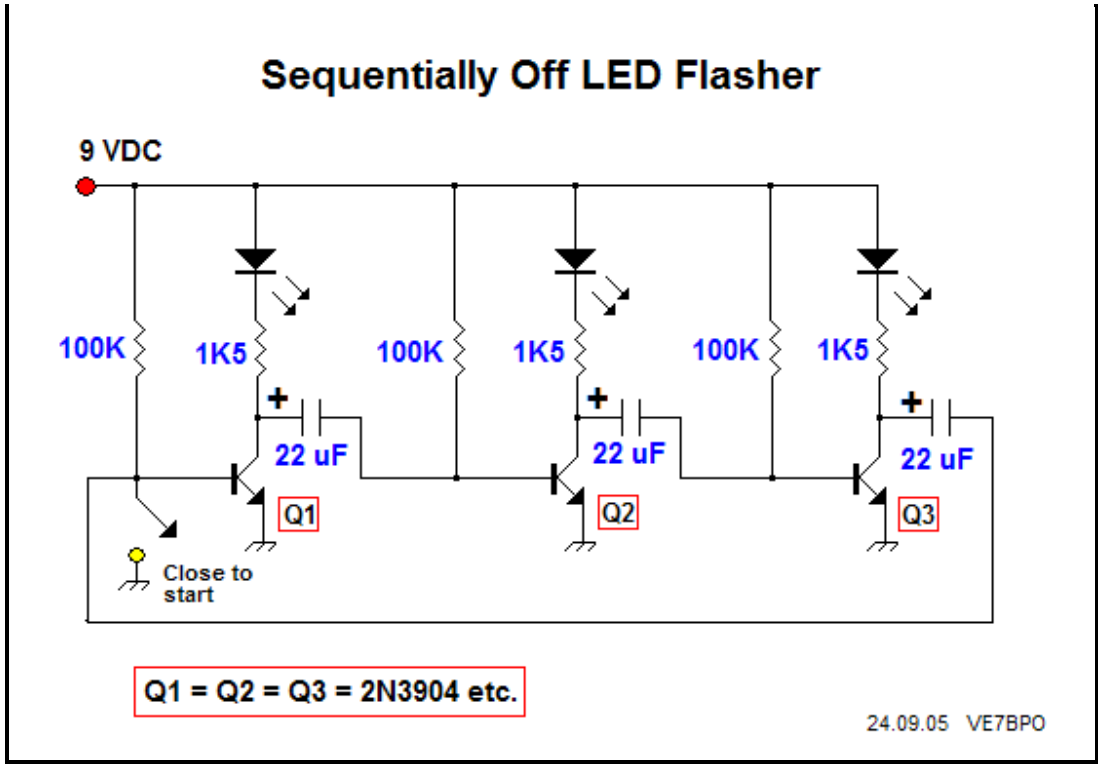
One of the four "nite-rider"

project chassis I built. After completion, this one was given to the son of VE7KPB. When drilling in a plastic chassis, I learned it is best to use a drill press set to a lower speed.

Sequentially Off LED Pulser

This circuit uses a series of transistors with an RC pair to pulse a string of LEDs.

This the favorite LED experiment I performed this summer. This flasher circuit is different in that it turns **off** alternate LEDs for about 1 second in sequence. When you connect this circuit to the 9 volt battery, all of the transistors are usually placed in saturation and therefore all the LEDs are **on**. Closing the switch on the base terminal of Q1 for a moment initiates the correct pulse sequence. The pulse initiates in Q1 which turns off the LED connected to the Q1 collector for about 1

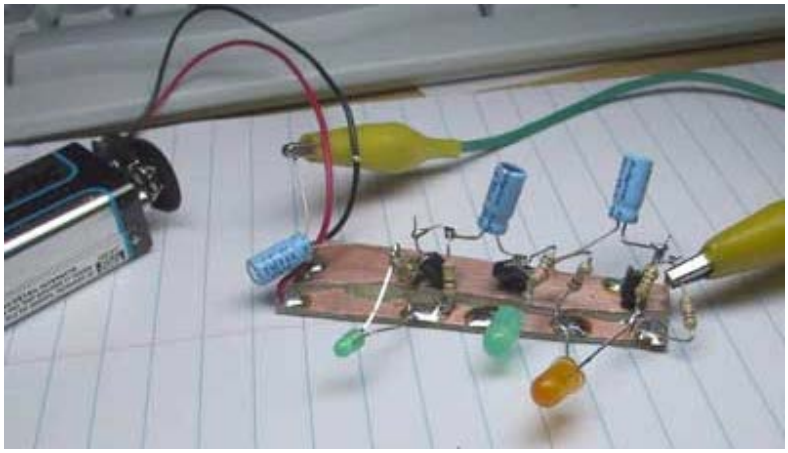


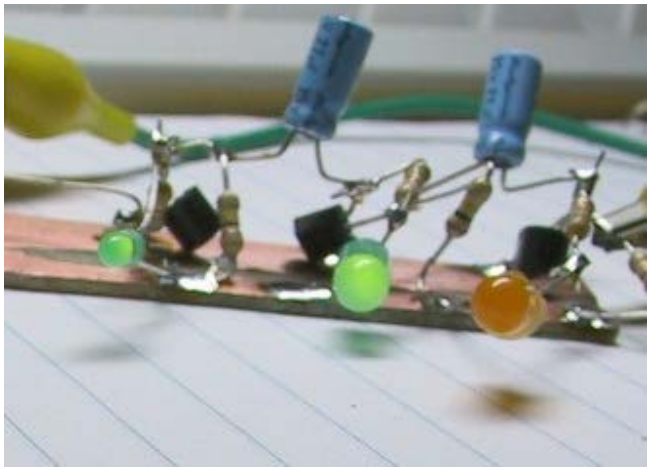
second. When Q1 turns back on (goes into saturation), Q2 turns off. When Q2 turns back on then Q3 turns off and so on. The circuit is a closed loop and many more stages may be added.

You can experiment with different base resistor and coupling capacitor values to vary the speed of the LED string or to create a sense of randomness by varying each transistor's RC stage separately. This is a fun circuit!

Youtube link (not mine).

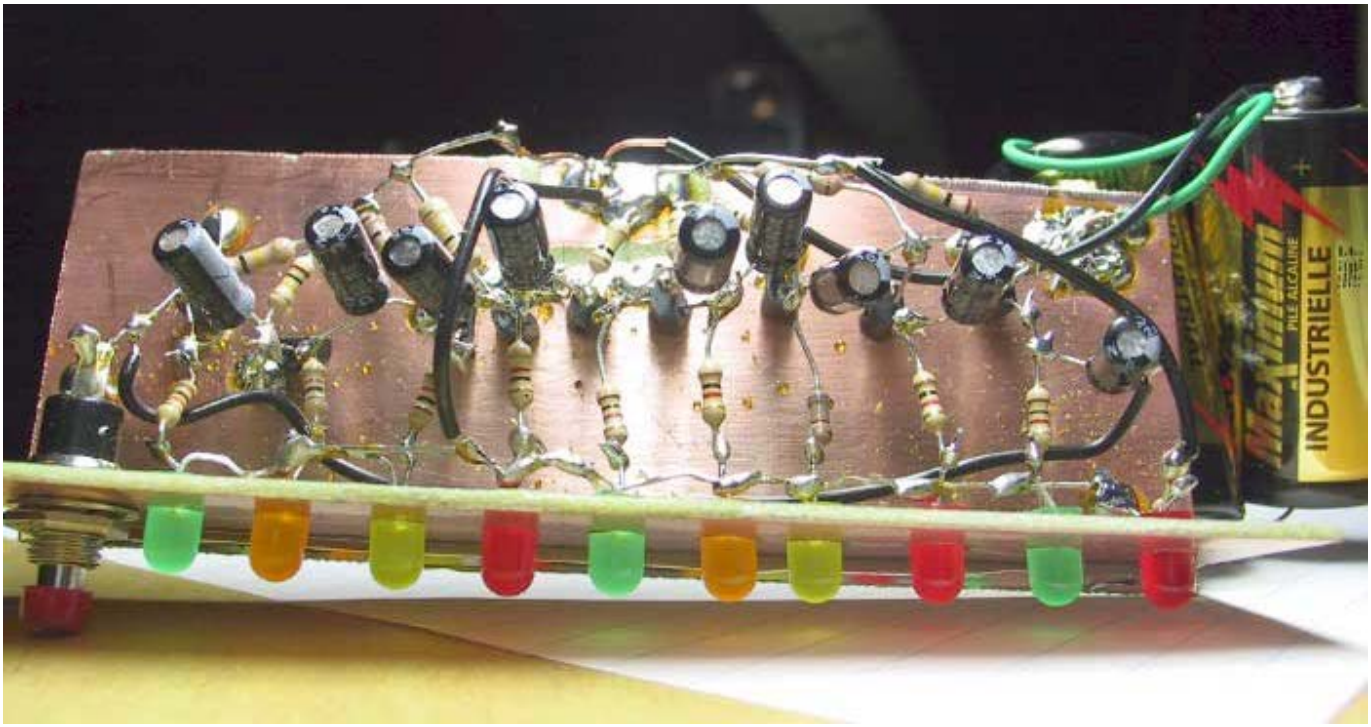
The prototype 3 transistor version. I just used a piece of wire to ground the Q1 base terminal and establish the correct pulse sequence after powering it up. For the LEDs, transistors, resistors and capacitors you can use whatever appropriate parts that you happen to have on hand. Current draw is less than 10 mA with a fresh 9 volt battery. Decrease the 1K5 current limiting resistor to 1K or so if you want brighter LEDs at the expense of more current draw. Do not operate this circuit above 9 volts unless you connect diodes from the transistor emitters to ground to prevent emitter-base breakdown.



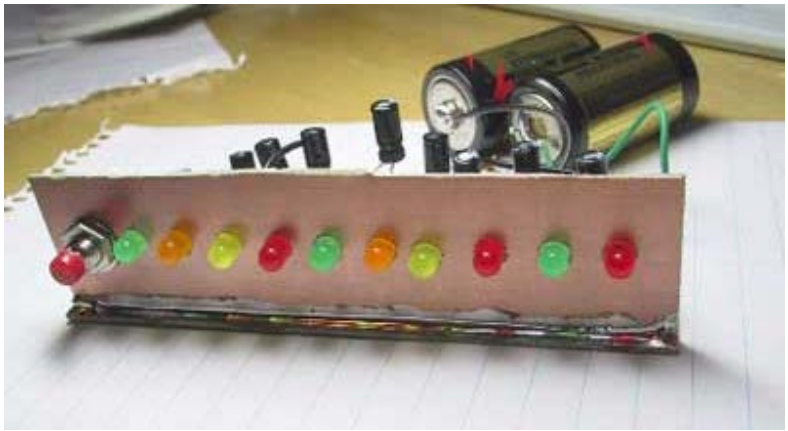


LED 1 and 2 are on and LED 3 is off at this moment in time.

Above and below photographs. This low current version has 9 LEDs connected in a chain and is powered by 3 volts. The 10th LED (extreme right hand side) is a flashing LED which is directly connected to the 3 volt supply and also uses a 1K current limiting resistor. Total peak current draw is only ~ 7 mA, yet it is still bright enough to see at night-time. The power supply is 2 D-cell batteries connected in series and then to the circuit by soldering wires directly onto the batteries with a 100 watt soldering iron.



NoNot counting the 10th flashing LED, 5 of the 9 LEDs are on at any given moment. A sequential flash effect is noted (the state of each LED flip-flops and shifts over 1 position each flash). If you build this project with an even number of LEDs, the sequential effect is not seen. Half of the LEDS (spaced every other LED) are on and the other half are off at any moment. The same LEDs are lit or unlit each pulse. Thus the effect is more like a typical multivibrator LED flasher. This variable, even versus odd number of stages property makes the circuit quite versatile.

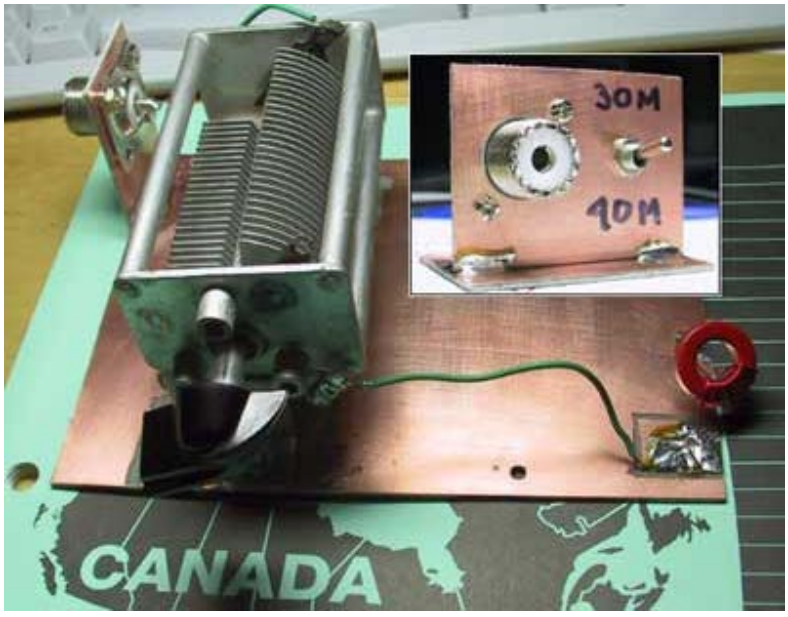


Conclusion

I hope that you have some fun experimenting with these and other circuits.

SWL Receiving Antenna Experiments

Introduction



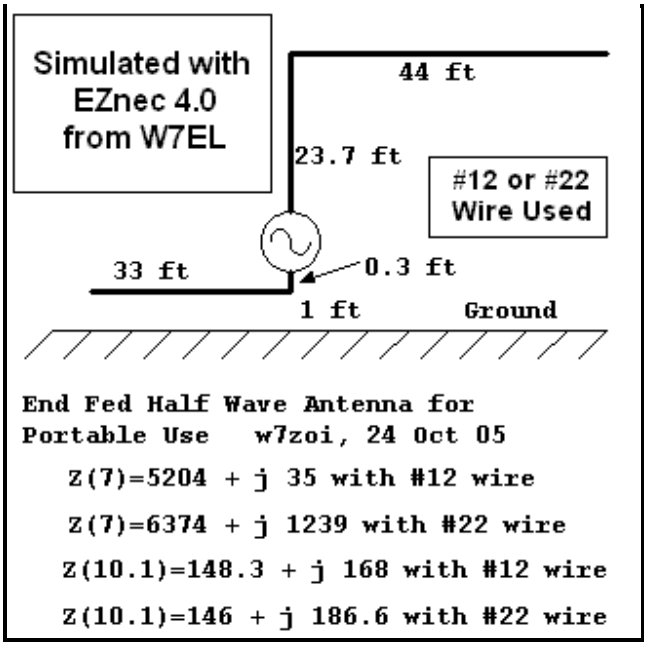
I have a lot to learn about SWL antennas. What follows are some brief experiments I performed in late October 2005. I have been experimenting with a half wavelength end-fed wire for use as a portable 40M band HAM transceiver (receive and transmit) antenna. This wire antenna is 67 feet long. End fed wires are very popular with those who pack a small portable transceiver when backpacking and camping. No feed line is required and the far end of the wire can be strung up using objects such as nearby trees or collapsible, portable poles. An elaborate ground system is not required. The return for the RF energy to ground might be grounding rod(s), short or long radial(s), or even just capacitively coupling to the local environment (including the operator!). Simple tuners are easily built to transform the high (thousands of ohms) wire impedance to the 50 ohms or so required by the transmitter.

I wanted to know if I could use this antenna as a tunable receive antenna for the the 30 and 31 Meter bands in addition to a tunable transmit/receiver antenna for my HAM radio work on 40 Meters. **What I verified is that tuning a multiband receive only antenna is not very practical.** When you tune a receive antenna you increase received noise and desired signals proportionally and therefore do not improve the signal to noise ratio in a meaningful way. Sometimes until you perform some experimentation, you don't really believe even good advice.

To the right: A computer simulation of a 40 Meter band end-fed Wire performed by W7ZOI on W7EL's EZnec program. The simulation was for 10.1 and 7.0 MHz with a 22 and 12 gauge wire. One 33 ft radial was used from the base of the 23.7 ft piece up 0.3 feet from the ground in this simulation. Z is impedance. Z and j are complex numbers used to represent the multi-dimensional quantities of the AC analysis of this antenna. In actual fact, j is an imaginary number. I suggest you might just ignore j unless your are well informed about impedance arithmetic.

Antenna Matching

A tuner can help match the impedance of the wire antenna and feed line (if used) to the input impedance of the receiver at a given frequency. This will result in more received signal and noise voltage to the receiver's input. HAM radio enthusiasts use antenna tuners to transform the impedance between the radio and the antenna tuner to 50 ohms to allow maximal output power from their transmitter. Non- amateur radio operators, can not use transmitters to



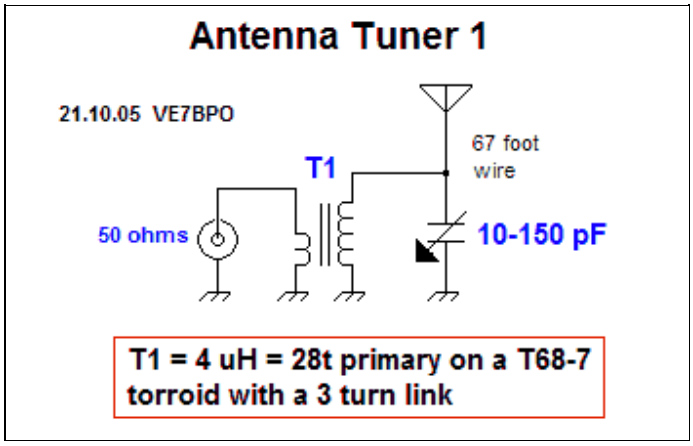
match their antennas. This leaves either using receiver noise, S-Meter or an antenna analyzer such as the MFJ259. I just used my ears and S-meter. All of the tuners presented work as transmitter tuners as well. Any network used for transmitter work must be able to handle the output power of the transmitter final amplifier.

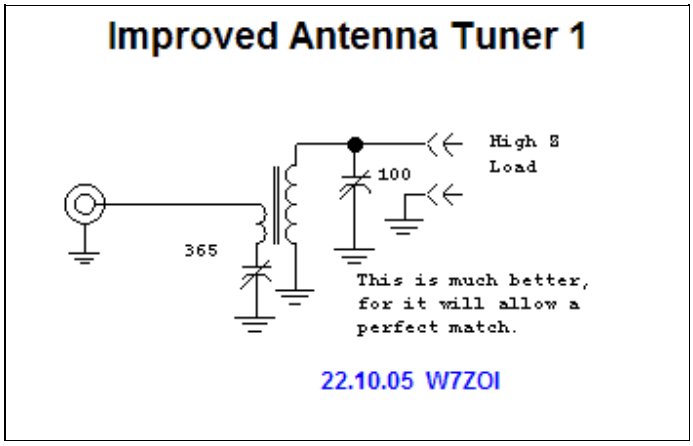
The Wire Antenna Experiments

I tried 3 different antenna tuners to see if I could tune an 18 gauge wire on 40-41 and 30-31 meters. My wire went from my computer room in the basement out a hole in the wall and sloped at ~ 50 degrees up to a rope tied to a tree in my backyard. The tip of the antenna is about 50 feet (15 meters) high. A 10 gauge insulated ground wire also passes from the computer room outside under the back lawn. It is "earthed" to two, 2 meter copper grounding rods hammered into the ground. No direct connection to the house ground system and the outside antenna grounding system should be made as this may result in increased receiver noise.

Antenna Tuner 1

The schematic on the left below, is very popular with HAMS who use it to to tune monoband end-fed wires. It is very simple and works reasonably well. Although the capacitor was able to resonate the 35 turns inductor, the T1 turns ratio was wrong and reducing it to 28:3 was required to get the maximum receiver signal in my experiments on the 40 and 41 meter bands. The alternative was to shorten or lengthen the wire antenna which is not very practical as it meant repeatedly climbing a tree.

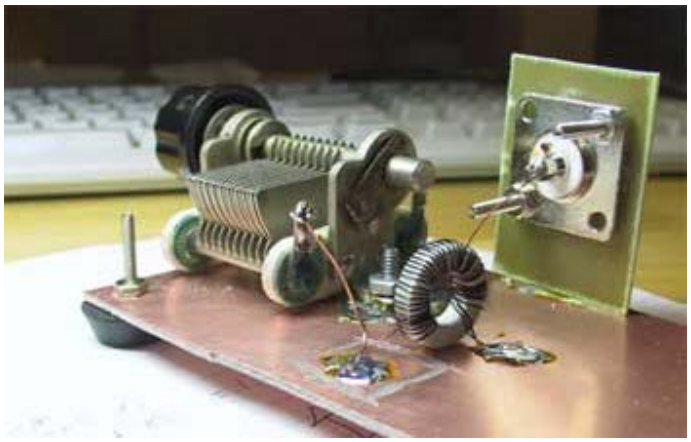


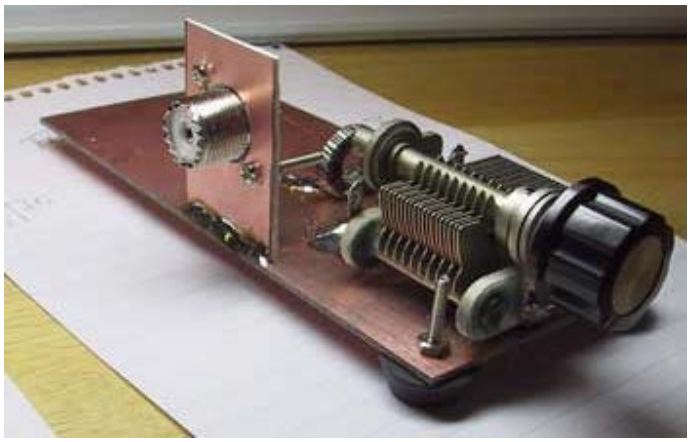


Above right: An improved version of the Antenna Tuner 1 schematic from W7ZOI. This tuner has 2 user "tweaking" adjustments much like most modern commercial antenna tuners (which typically also have a band changing adjustment).
I did not build this version.

In addition, this tuner would not tune the 67 foot wire on the 30 or 31 meters band. This was no surprise. When I turned the variable capacitor, I noticed some change in received signal, but not much. The signal strength was very poor and I could not hear much of anything.

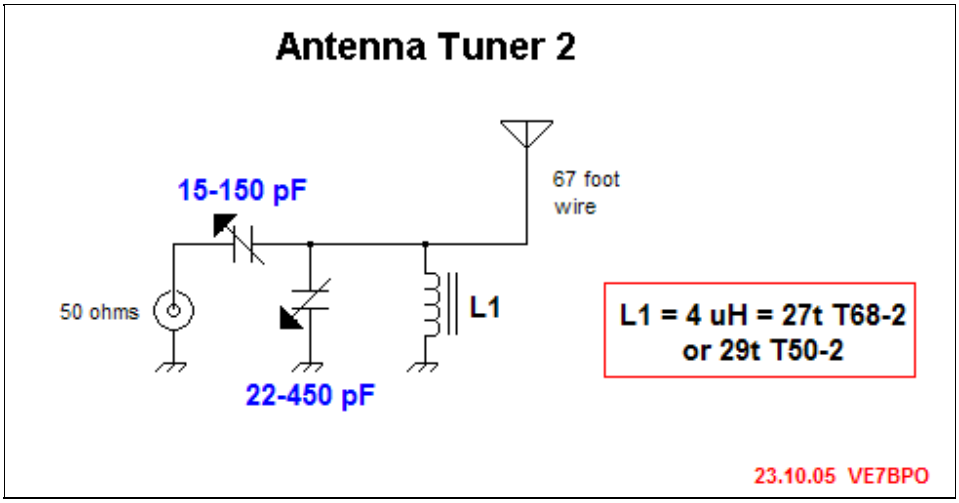
Below. Two built up views of the Antenna Tuner 1 schematic. I soldered the antenna wire to the circuit in the isolated area connecting the inductor and capacitor. The outside antenna ground wire was soldered to the large copper ground plane.

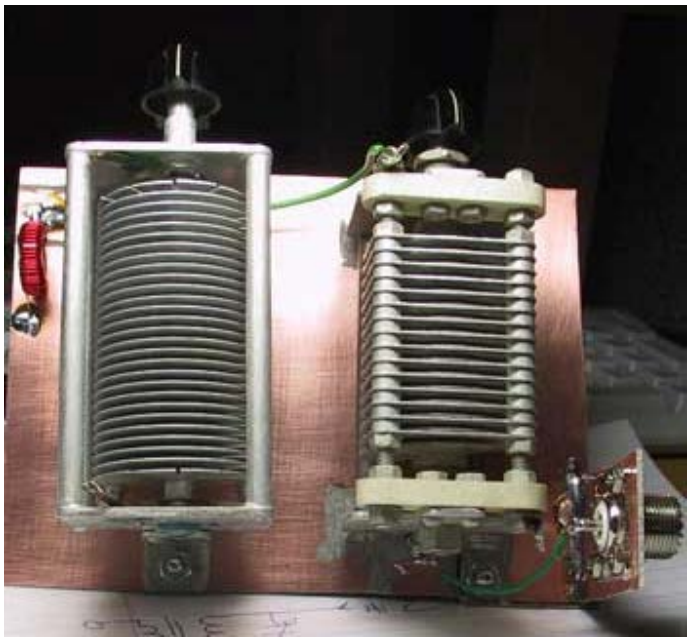


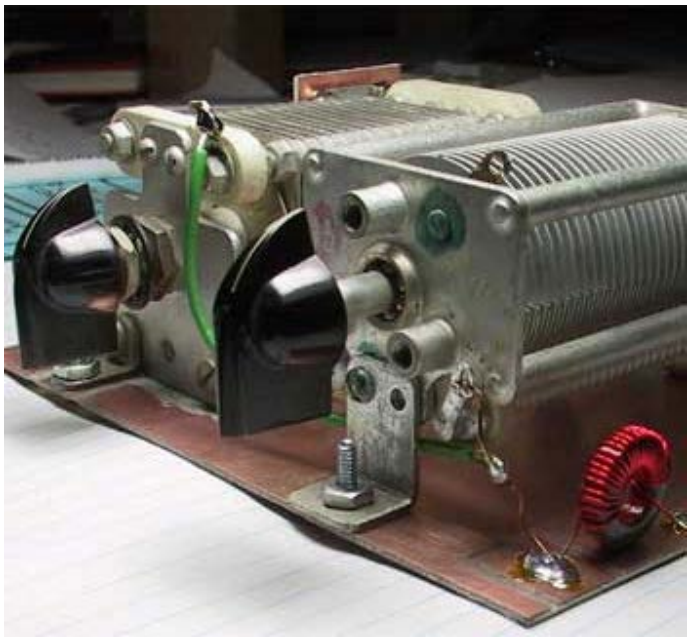


Antenna Tuner 2

This was the next tuner I built. Antenna tuner 2 tuned very sharply on the 40 and 41 Meter bands. It is designed to match a high impedance antenna, so it could not match the medium impedance (~150 ohms) wire to the 30 and 31 meter bands very well at all. Since ~ 150 ohms is fairly close to my receivers 50 ohm input impedance, I just connected the wire antenna directly to my receiver. The received noise and signals were then much stronger than those with the Antenna Tuner 2 network in the circuit on 30-31 meters.



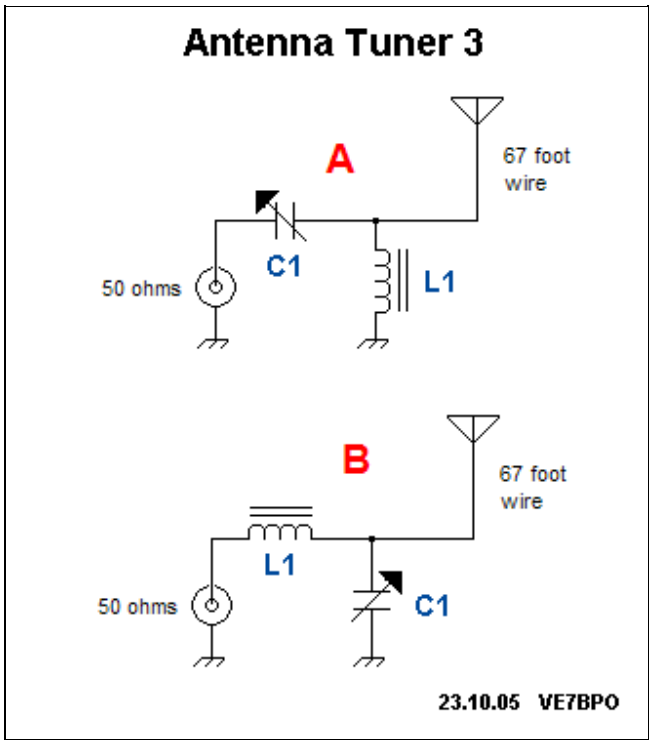


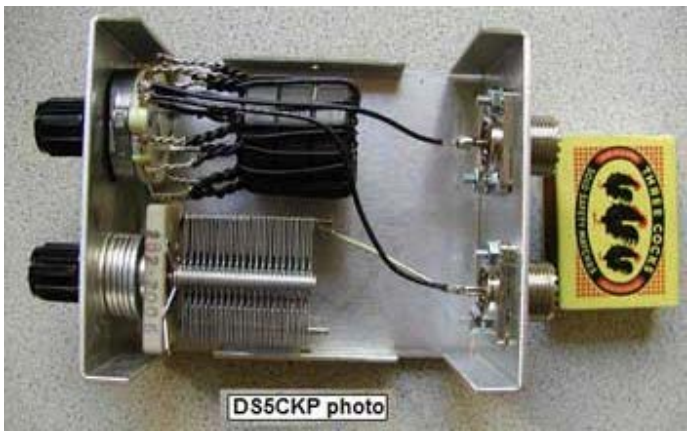


Above: Two constructed views of the Antenna Tuner 2 schematic. The variable capacitors were bought at a HAM festival in 1992 for 2 dollars each.

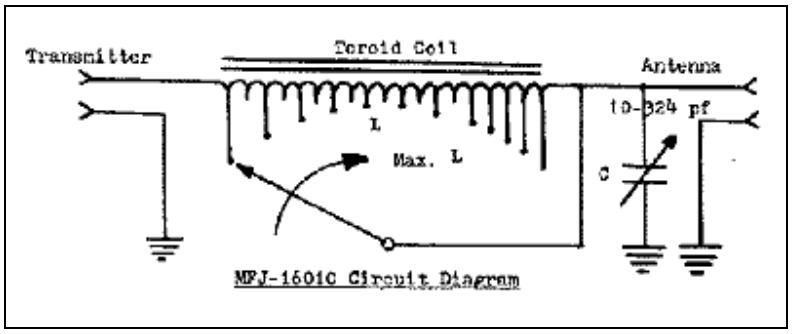
Antenna Tuner 3

The next antenna tuner topology I tried was the familiar L network. Circuit **A** is configured in a shunt L (inductor to ground) and series C (capacitor) and is a high pass L network. Circuit **B** is configured in a series L and shunt C and is a low pass L network.





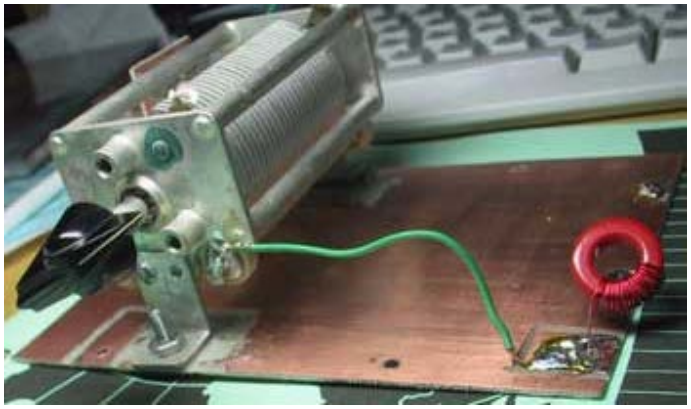
L networks (especially the low pass form) are very popular as random wire tuners. MFJ sells an excellent version as the model MFJ-16010. A photograph of this tuner is shown below right. The above left photograph was taken by **DS5CKP** who also sells a random wire antenna tuner at: http://user.chollian.net/~cyberline/ckptuner.htm

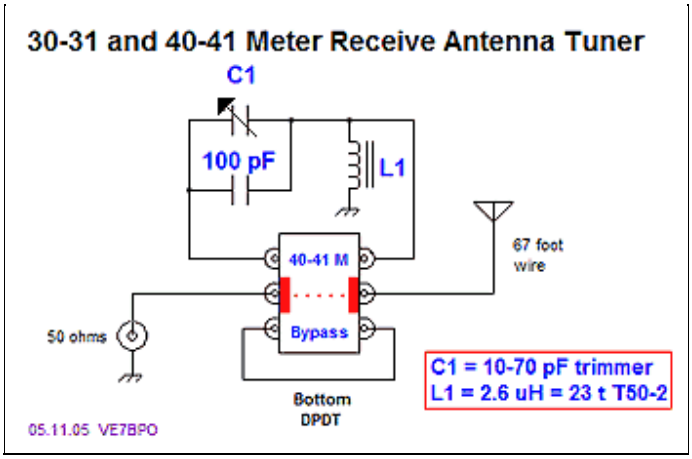


Above. MFJ web publishes their manual including the schematic for the MFJ-16010 random wire antenna tuner. The inductor is actually wound on 3 stacked (probably ferrite) cores which are tapped. The taps are connected to a front panel mounted 12 position switch. This allows coverage from 2-30 MHz. As a simple experiment, I tried stacking 2 and then 3 FT-50-61 ferrite cores and was able to get a wind range of inductances from the 6 taps I made on my test inductor.

Numerous examples of the L network antenna tuner can be found on the web and in print including the 2006 A.R.R.L. handbook. I decided to try the high pass L network topology to experiment with.

I built part **A** of the Antenna Tuner 3 schematic with shunt L and series C. On the 40-41 meter bands I tried 3 different inductors; 4 uH, 2.1 uH and 1.3 uH. I tuned the network to get the greatest receive noise and S meter reading and measured the variable capacitor value. At 7.30 MHz, the capacitance values were 86 pF, 141 pF and 182 pF respectively. Although non-critical, I settled on a 2.6 uH inductor (23 turns on a T50-2 powdered iron torroid) so I could use a junk-box 10-150 pF variable capacitor to resonate it. Later I decided just a trimmer capacitor might do. You need around 125 pF at 7.30 MHz to tune the network with a 2.6 uH inductor (to give you a ballpark C value to start with).

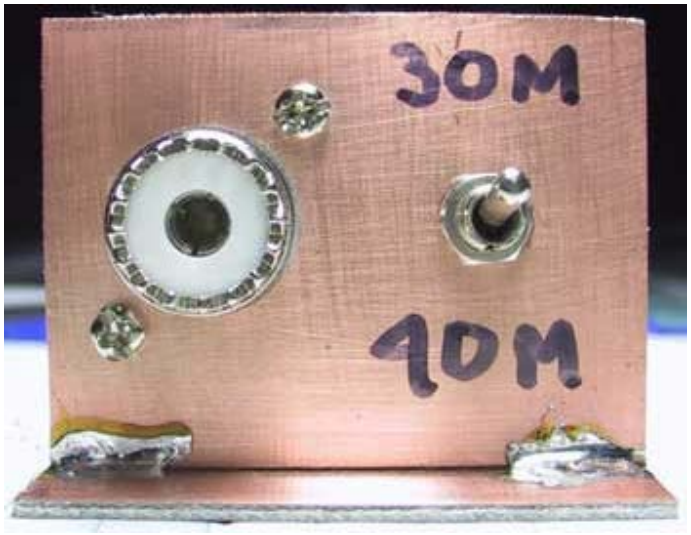


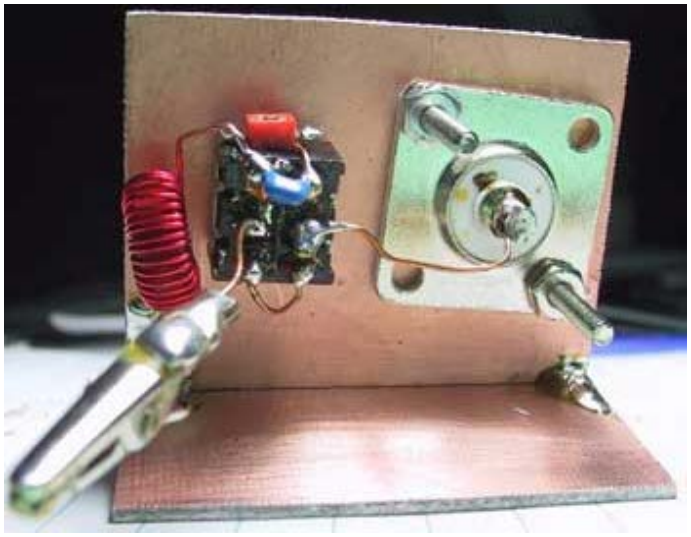


Schematic to the right: A very small receiver tuner that allowed the L network to be switched in and out of the antenna path was constructed. The L network is tuned on 40-41 meters via a small trimmer cap on top of the double pole, double throw switch. A small plastic alignment screw driver is used for signal peaking. An air variable capacitor would be much easier to tune as you move up and down these bands.

Below left: Front view. Technically, the switch label should indicate **40M** and **bypass** as the bypassed antenna could be used on any band.

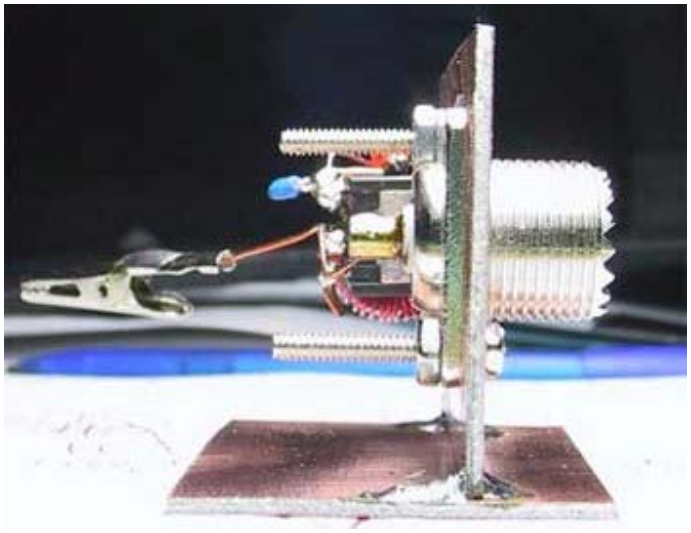
Below right: Rear view The trimmer cap can be seen on top of the switch.





Below left: Side view, the antenna wire is held via an alligator clip. This allows me to unclip and ground the antenna when it is not in use. The antenna ground wire is soldered to the copper ground plane.

Below right: A T6-8 core wound with with 24 gauge wire. I used larger powered iron torroids to wind my inductors, however, they can also be wound on ferrite cores or be air-wound. The core size, wire gauge and type of inductor used can affect the selectivity and insertion loss of the tuner network, however for practical receiving purposes it is not a major concern.



Conclusion

The final experiment allow me to easily switch between a matched and direct 67 foot wire antenna on 40 and 41 meters. I performed several listening tests and generally agree with those who say receive antenna tuners offer little to no improvement in signal readability. The only advantages I can think of for matching a receive antenna to a reasonably quiet and sensitive receiver are:

1. There may be some improvement in the receiver front end filter function as these filters are designed to have a specific input impedance.
2. Certain balanced mixers may function better with the correct impedance on their RF port.
3. The tuner itself (depending on design) may marginally improve the front end selectivity of the receiver it is connected to.

I do like the noise roar and louder signals with the matched antenna on 40 and 41 meters, although this is totally subjective. I think the reason for this is that as a HAM radio operator who always matches the antenna for any band I am on, I am used to louder signals and noise levels. I also spent most of my first 10 years of HAM radio operation on 80 and 160 meters which are relatively noisy bands and have been conditioned so that noise is "normal".

On 30-31 meters the bypassed antenna worked quite well and the L network can be switched in as an attenuator. Perhaps you might build up a tuner and try for yourself!

Although they are simple, low cost and easy to put up, it is likely unwise to use an end-fed wire without coaxial feed line as a receiving antenna. The ground wire is part of the antenna system and easily picks up household generated noise which will present to your receivers input. The time honored and easiest methods to reduce receiver noise are to get your antenna away from the house and other noise sources, use buried coaxial cable feed line to the house and directly earth ground the shield of the feed line with stakes. Antenna/ feed line "link coupling" by a transformer may also reduce noise (especially if the antenna system is balanced) and this topic begs further study.

Suggested Links

There are some fabulous web sites on the topic of SWL antennas and reducing receiver noise. I suggest these 5. Try a Google search for more.

http://www.dxing.info/equipment/

http://www.hard-core-dx.com/nordicdx/antenna/feed/feed1.html

http://www.aa5tb.com/efha.html

http://www.nyx.net/~dgrunber/

As usual, I learned a great deal from the process of experimenting. I look forward to spring when some more antenna experiments can be performed.

Experiments with JFET Biasing

The most common way of biasing a Junction Field Effect Transistor (JFET) is with a source resistor. This method, shown in Fig 1 below, has the advantage of offering negative feedback to stabilize the bias conditions. This is the same thing that happens when a bipolar transistor uses an emitter resistor. Self bias can be used as a method to evaluate a JFET to determine the critical parameters that describe it: Idss and Vp. These are discussed in Chapter 2 of **Experimental Methods in RF Design** and many other places.

The method used in our experiment is to set up the FET of interest in a test fixture with a power supply, bypass capacitor, resistors in the drain and gate to suppress parasitic oscillations, and a handful of extra resistors, R-test, that can be paralleled with an existing 100K source resistor. A digital volt meter (DVM) is the basis for the measurements. We begin by using the DVM to measure the resistance of our test resistors, for the values will be used in calculations. The DVM is then attached to the FET source to measure the DC voltage. The first value we measure is with no attached R-test. The measured value will be very close to the FET pinchoff voltage.

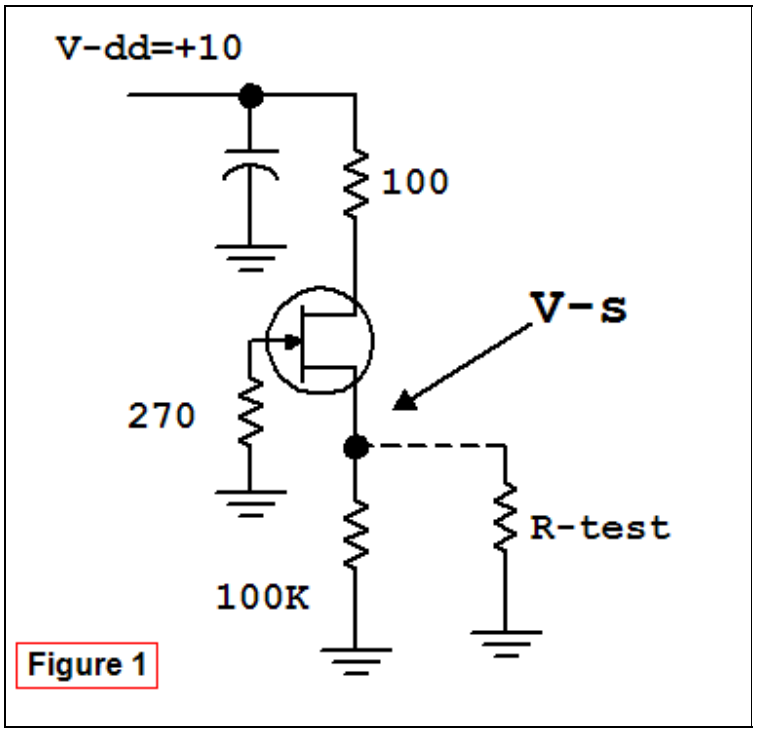


Figure 1

The measurements we will perform infer drain current as a function of gate to source voltage. The physics of the FET support the model that there is no gate current so long as the gate is not forward biased with regard to the source. Hence, the drain current equals that in the source. We will measure the source current by measuring the voltage drop across the source resistor. The gate is at ground potential, for there is no gate current, so the gate to source voltage is just the negative of the source to ground voltage.

The resistors that I pulled from my stock for some measurements were marked as 22, 39, 68, 100, 150, 300, 510, 680, 1K, 2K, 3.3K, 6.8K, and 10K Ohm. The measured values are shown in attached figures. A systematic pattern was noticed with all of the measured resistances under the marked value, suggesting an error in the calibration of my DVM, a Fluke Model 73. All resistors were 2% carbon film 0.25 Watt. However, when I measured a 499 Ohm, 1% metal film resistor, it came up exactly at 500 Ohms. The differences between the measured values and those marked on the part were small enough that I neglected the details and used measured values for calculations.

The first FET I examined was a 2N5454, a common JFET that I had in my junk box. The source voltage was 3.26 with nothing but the 100K for source bias. I started my measurements with the largest resistor, 10K. The voltage dropped to 2.90 and was stable. I merely held the resistor in place rather than soldering it. The resistor was kept in place long enough to get a stable reading that I could record in my lab notebook. All results were of the same character until I got to the 300 Ohm resistor. At that point I started to notice a slight heating effect. The source voltage was 1.523, but slowly dropped to 1.518 volts. This behavior continued through the lower value resistors. The 22 Ohms produced 264 mV on the source that then dropped to 256 mV.

Later I examined a J310 JFET. This is a much larger area part than the 2N5454 with an Idss that is about three times larger. With the 22 Ohms in the test fixture, V-source went to 701 mV, but settled at 652 mV. The drain current was then 32 mA. With a 10 volt power supply, there was nearly 200 mW dissipated in the FET. This is within ratings, but high enough to produce heating. Operation at higher voltages and at Idss would further tax the part. One must take care when doing these measurements to be sure that the source voltage is observed quickly.

Attached are the MathCad documents that I used to examine the data. A spread sheet such as Excel could be used, but I prefer the graphics of MathCad. The second page for the 2N5454 shows a graph for the observed data as well as a calculated one. The two FET parameters for the 2N5454 were varied to obtain a good correspondence between the two. The part had Idss=15 mA with Vp=-3.5 volts. This is similar to the popular MPF-102, but close to the high Idss extreme for that part.

The data presented for the J310 is more abbreviated with only two points shown. I picked the 22 Ohm and 1K source resistors. This still produced data that is very close to that obtained with many more data points.

My initial analysis suggested that we could characterize the FET by measuring the source voltage with 100K in place to approximately determine Vp, and to then short circuit the source through the mA scale on the DVM to obtain Idss. This is a reasonable start. However, the pinchoff will usually be a few percent more negative (for an N-channel depletion mode part). The long leads in the source also make me feel uncomfortable with regard to parasitic oscillations.

After the DC measurements were done, I thought it wise to look at the potential for oscillation. The J310 was in the test circuit at this time. The TO-92 J310s are parts that are well known for their propensity for oscillation, so I guessed that it would not be difficult to coax this one into such a mode. But this was not what I found. I eliminated the 100 Ohm drain resistor, but moved the FET close to the 0.1 uF bypass. This bypass is not a very good one for VHF and upward. No oscillation was seen. I then eliminated the gate resistor, replacing it with the normal gate lead. Still no oscillation. I eventually added a gate inductor and a parallel tuned circuit in the source. The source bias resistor had a RF choke in series with it. I finally saw a robust VHF oscillation, but nothing else up through 1.5 GHz.

Experiments with JFET Biasing, after correspondence with VE7BPO on 14Feb06. (See earlier Mathcad file "Jfet_bias1_ve7bpo.mcd) Measurements on 2N5454.

$n := 0..12$ **index for measurements and calculations. Vsn=measured source voltages.**

$$Rx_n := \begin{pmatrix} 21.6 \\ 38.6 \\ 66.9 \\ 98.3 \\ 149.4 \\ 298.3 \\ 508 \\ 671 \\ 998 \\ 1997 \\ 3270 \\ 6720 \\ 9810 \end{pmatrix}$$

<== Measured Rx values

R values paralleled with 100K ======>

$$Rp_n := \frac{1}{\left(\frac{1}{Rx_n} + \frac{1}{100400}\right)}$$

$Rp_n =$

	0
0	21.595
1	38.585
2	66.855
3	98.204
4	149.178
5	297.416
6	505.443
7	666.545
8	988.177
9	$1.958 \cdot 10^3$
10	$3.167 \cdot 10^3$
11	$6.298 \cdot 10^3$
12	$8.937 \cdot 10^3$

$$Vs_n := \begin{pmatrix} 0.264 \\ 0.429 \\ 0.654 \\ 0.85 \\ 1.09 \\ 1.523 \\ 1.845 \\ 2.001 \\ 2.203 \\ 2.492 \\ 2.652 \\ 2.82 \\ 2.90 \end{pmatrix}$$

$Vg_n := 0 - Vs_n$

$Vg_n =$

	0
0	-0.264
1	-0.429
2	-0.654
3	-0.85
4	-1.09
5	-1.523
6	-1.845
7	-2.001
8	-2.203
9	-2.492
10	-2.652
11	-2.82
12	-2.9

Use Mathcad "vectorize" operator.

$$Id_n := \frac{\overrightarrow{Vs_n}}{Rp_n} 1000$$

$Id_n =$

	0
0	12.225
1	11.118
2	9.782
3	8.655
4	7.307
5	5.121
6	3.65
7	3.002
8	2.229
9	1.273
10	0.837
11	0.448
12	0.325

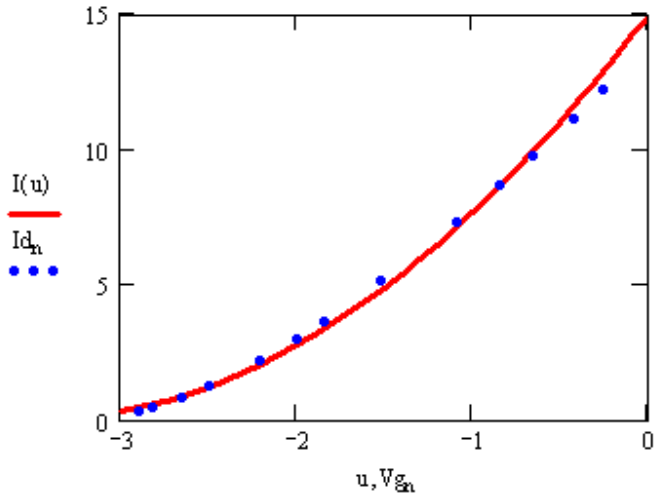
Let pinchoff be $u_p := -3.5$ **and** $I_{dss} := 15$

$u := -3, -2.9 .. 0$

$$I(u) := I_{dss} \cdot \left(1 - \frac{u}{u_p}\right)^2$$

Try various values for pinchoff and Idss to get a reasonable match.

Blue is measured data while red shows calculated behavior.



Experimental Note: During the measurements, I noticed a definite thermal change in Idss when dissipation was high. This occurred at higher currents, with low source bias resistors. The Vg values I used in my data were the first ones recorded with the DVM. The most extreme case was with the 22 Ohm source R. The source V started at 264 mV and went down to 256 mV. The effect was more dramatic with a larger part, a J310. That part, when operated with a 22 Ohm R, started with 701 mV and dropped to 652 mV.

Experiments with JFET Biasing, after correspondence with VE7BPO on 14Feb06. (See earlier Mathcad file "Jfet bias 18feb06.mcd) Measurements on J310.

Here we examine a minimal data set of just two points.

$n := 0..1$ **index for measurements and calculations.**

$Rx_n := \begin{pmatrix} 21.6 \\ 998 \end{pmatrix}$ **<== Measured Rx values.**

$Rp_n := \dfrac{1}{\left(\dfrac{1}{Rx_n} + \dfrac{1}{100400}\right)}$

Vsn=measured source voltages.

R values paralleled with 100K ======>

$Rp_n = \begin{pmatrix} 21.595 \\ 988.177 \end{pmatrix}$

$Vs_n := \begin{pmatrix} 0.701 \\ 2.97 \end{pmatrix}$ $Vg_n := 0 - Vs_n$

$Vg_n = \begin{pmatrix} -0.701 \\ -2.97 \end{pmatrix}$

Use Mathcad "vectorize" operator.

$Id_n := \dfrac{\overrightarrow{Vs_n}}{Rp_n} 1000$ $Id_n = \begin{pmatrix} 32.461 \\ 3.006 \end{pmatrix}$

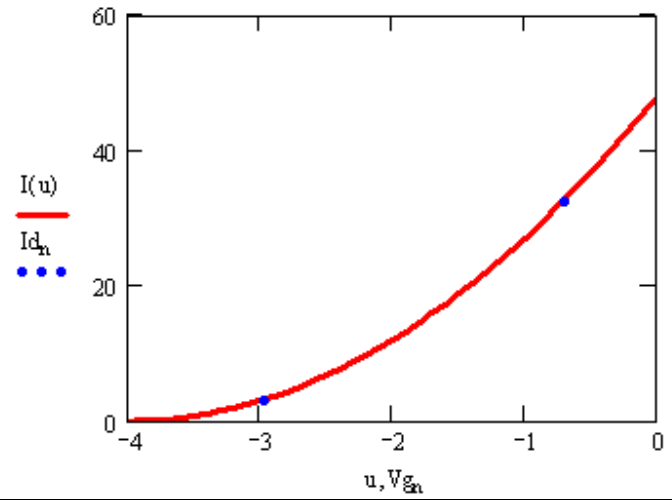
Let pinchoff be $u_p := -4$ and $I_{dss} := 48$

$u := -4, -3.9..0$

$I(u) := I_{dss} \cdot \left(1 - \dfrac{u}{u_p}\right)^2$

Try various values for pinchoff and Idss to get a reasonable match.

Blue is data while red shows calculated behavior.



18 Feb, 2006 by W7ZOI

Many thanks to Wes, W7ZOI for this contribution.

10 MHz WWV Receiver Experiments

New feature: Click on a schematic to load it into a separate browser page for printing

Introduction

For nearly a year, I have been trying to develop a tuned radio frequency (TRF) 10 MHz, WWV, AM receiver. My initial RF stages were common emitter or common source stages with tuned input and output. Despite careful layout, parasitic oscillations plagued these designs and they were discarded. Later, I discovered that only tuning the input of RF stages reduced this tendency towards instability and still provided reasonable selectivity. Different detectors were also tried and evaluated.

A simple receiver that sounds great and is fun to build and experiment with follows. My special thanks to Wes, W7ZOI for performing many of the simulations and providing suggestions which kept me going.

WWV Audio Files from Sept 26, 2006

WWV File 1

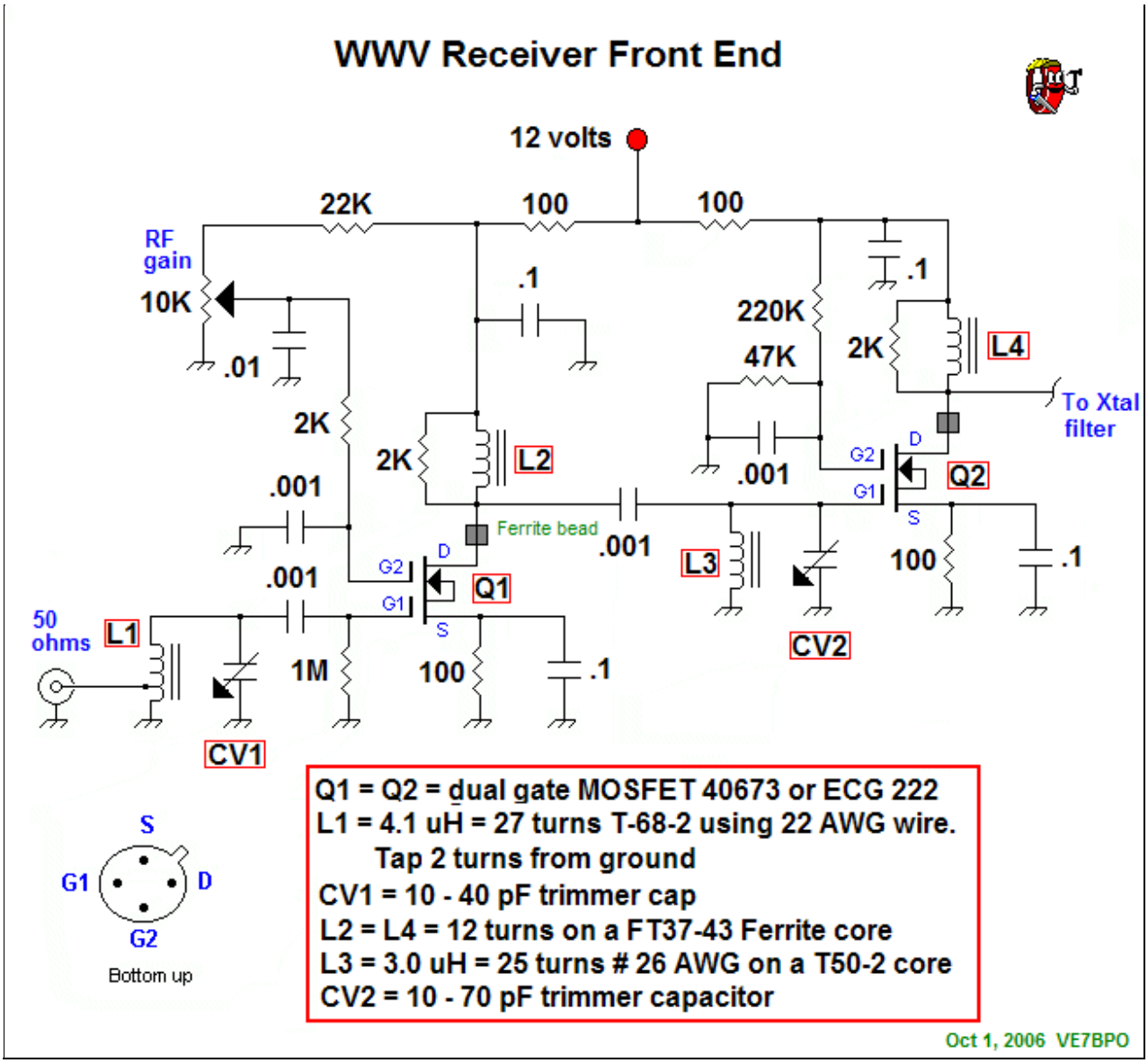
WWV File 2

WWV File 3

The audio was digitally recorded using an electret condenser microphone held 3 cm away from the receiver speaker. The files were compressed using the WMA format.

Supplemental Web Page added June 29, 2007

Receiver Front End



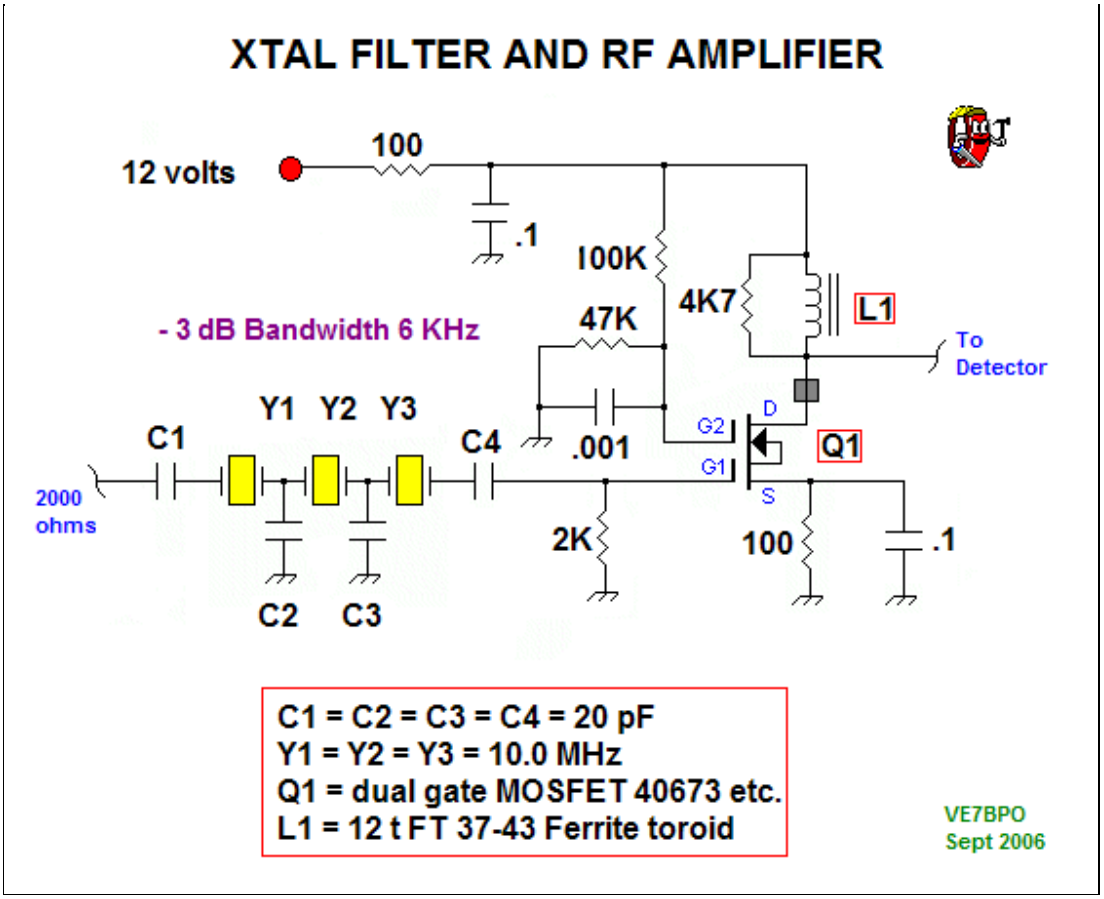
Above schematic. The receiver front end has just 1 single-pole filter. For even greater selectivity (but greater insertion loss), consider moving the L1 tap to 1 turn from ground. My receiver was connected to a 80 meter dipole via an antenna tuner. The antenna tuner provided additional selectivity. No local broadcast band (BCB) signals were heard when the chassis lid was tightened on. You may require additional RF high pass or band pass filtering in your location.

The RF gain control is very basic and only the first 1/3 of the 10K pot is used to go from minimal to maximal gain. Modifications to allow more precise variation in RF gain for dual gate MOSFETs or cascode JFETS are shown in EMRFD. The method shown works fine. For the most part, I keep it set to minimal gain. Using higher gain than necessary, increases receiver noise and may overdrive the detector.

TRF receivers require careful layout. A piece of wire greater than 2-3 cm between the stages may be enough to plague your receiver with local BCB interference depending on your layout and chassis integrity. For interstage connections that had greater than a 2 cm gap, shielded 50 ohm cable was used to prevent BCB interference.

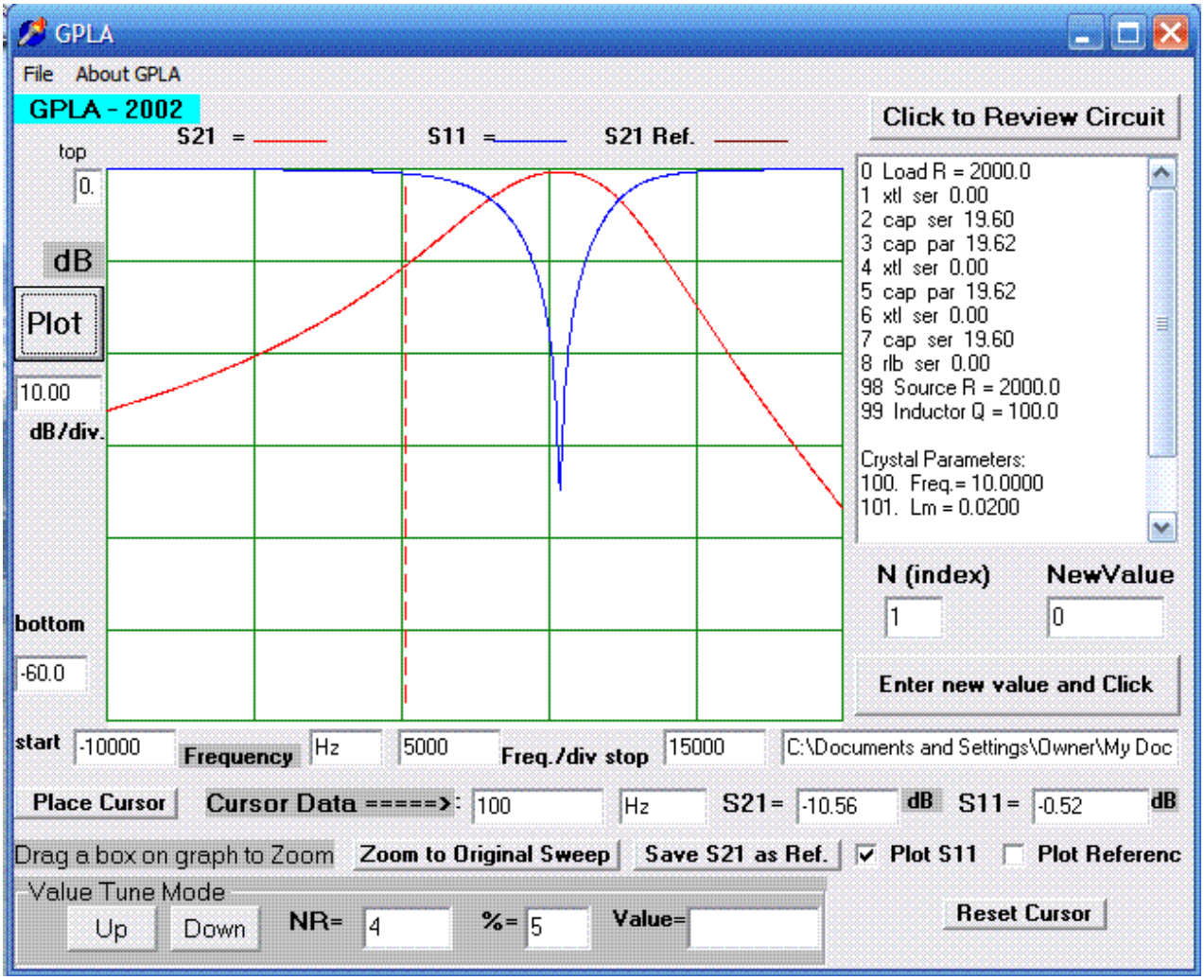
Dual gate MOSFETs provide adequate gain and low noise. you might consider cascoding 2 JFETS for each RF stage if you cannot obtain them. titleernatively, bipolar feedback amplifiers may be used and examples are provided later on this web page.

XTal Filter and RF Amplifier



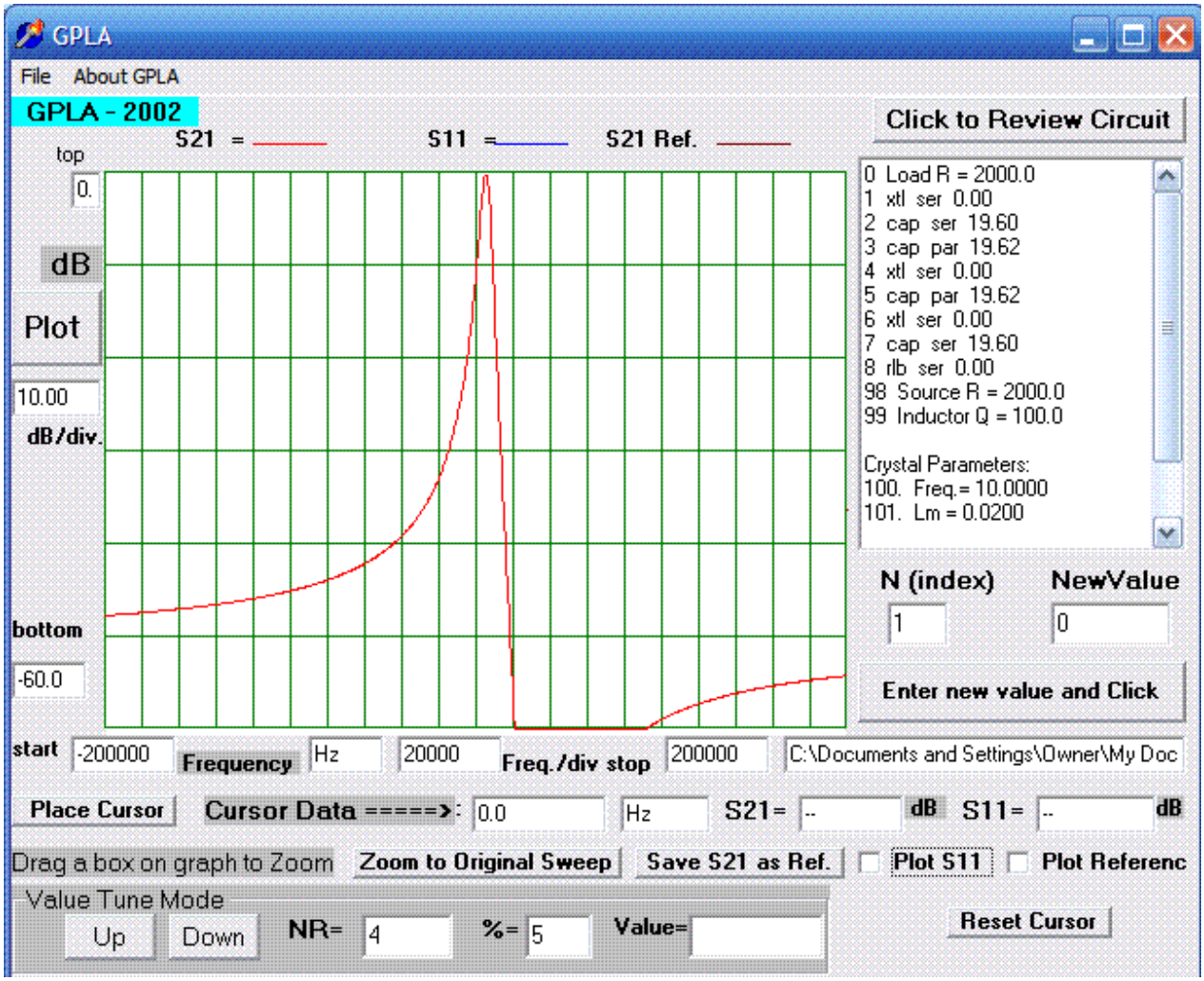
Schematic: The Q2 output impedance is 2000 ohms to match the input impedance of the Cohn crystal filter. This filter was designed by Wes, W7ZOI. Matched, computer grade, 10 MHz crystals were used. Choose 10 MHz crystals that are marked for a 20 pF or 32 pF load capacitance if possible. Using a 10 MHz crystal oscillator, find 3 that are closest to one another in frequency.

You may substitute 2K2 resistors instead of the specified 2K with a slight penalty in pass band shape.

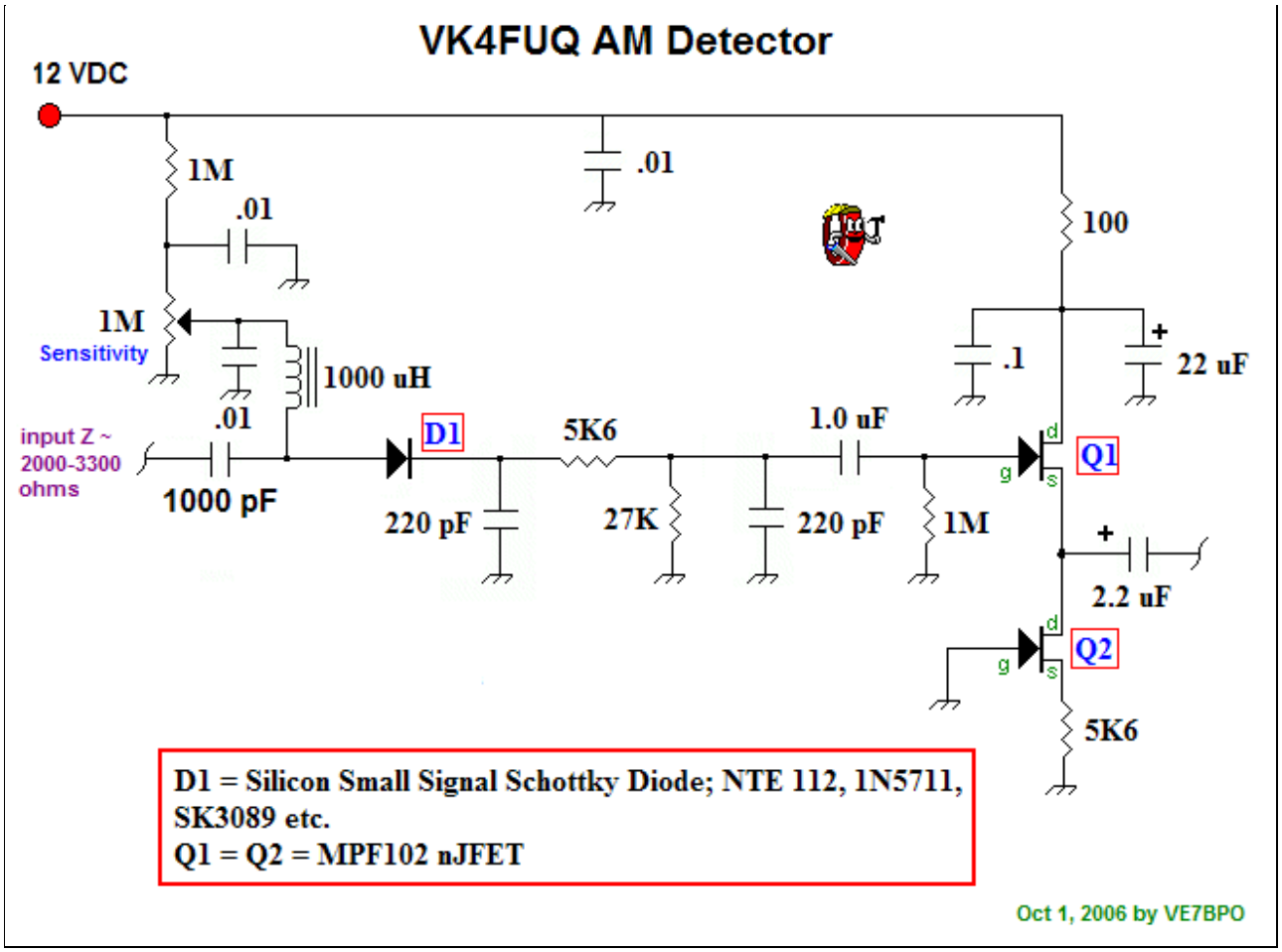


Above graphic. A simulation of the receiver crystal filter using GPLA, a program written by Wes, W7ZOI that comes with EMRFD. EMRFD is the major reference for this web site and I recommend that you add this book and companion software to your home library. The pass band is not symmetrical. It is mistuned for the lower pass band frequencies and would serve better as an upper sideband filter. Nevertheless, it works reasonably well and is simple to build and tolerant to component variation and match. At certain times, a very strong shortwave station at 9.985 MHz can be heard along with WWV. This usually occurs in the early evening when the WWV signal is not that strong at my location. For most of the day and night, whether WWV is present or not, very little interference has been detected.

Bypassing the crystal filter is an interesting experiment. As many as 5 stations were heard simultaneously and these varied as time passed. I heard Radio Vatican, Radio Habana and many other broadcasts during 1 evening. At one point I heard a station at 9.75 MHZ, WWV and a strong CW carrier at 10 .110 MHz!



Above graphic. This is a sweep that goes from 200 kHz below to 200 kHz above 10 MHz to show the stop band response of this filter. This filter has a pretty decent response considering the low cost and effort involved.



Above schematic. This detector is fabulous. It was designed by Felix Scerri, VK4FUQ. He has a web page explaining his high fidelity detectors at the Elliot Sound Products (ESP) site: http://sound.westhost.com/articles/am-radio.htm

The ESP web site is a personal favorite. Rod Elliot has one of the best do-it-yourself electronic web sites available. The main URL for his site is http://sound.westhost.com/index2.html

My sincere thanks to Rod and Felix for permission to present Felix's detector on this web page.

His improved AM detector has 3 positive advantages; it has high bandwidth, low distortion and incredible (and variable) sensitivity. I cannot get over how nice this detector sounds compared to others I have built and analyzed during weak and strong signal testing. The variable bias control allows the listener to adjust the bias to maintain detected audio fidelity even when the RF signal is weak.

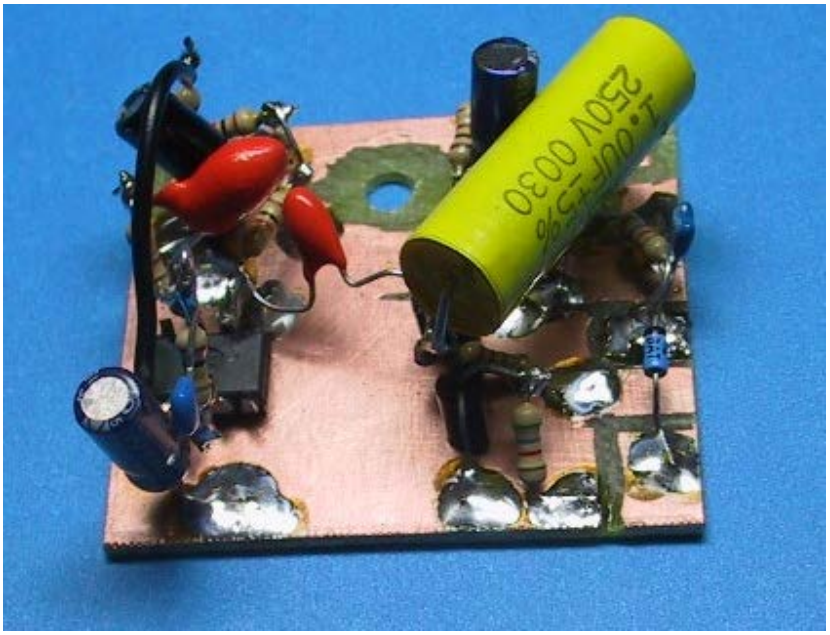
This detector uses a UHF mixer diode often found in older television sets. Increasing the diode bias from O volts towards maximum causes three things to happen:

1. Increased sensitivity.
2. Increased audio high frequency response.
3. Slight increase in receiver noise.

When the WWV RF signal is weak, turning the bias off may result in the detected WWV signal disappearing. Increasing the bias will bring WWV back in. I generally run the bias control pot about 1/2 way and of course, higher as WWV fades out. I like the fidelity that the bias adds even when the WWV signal is strong. Note how the WWV audio quality continues to be high in fidelity as WWV fades out in this sound file.

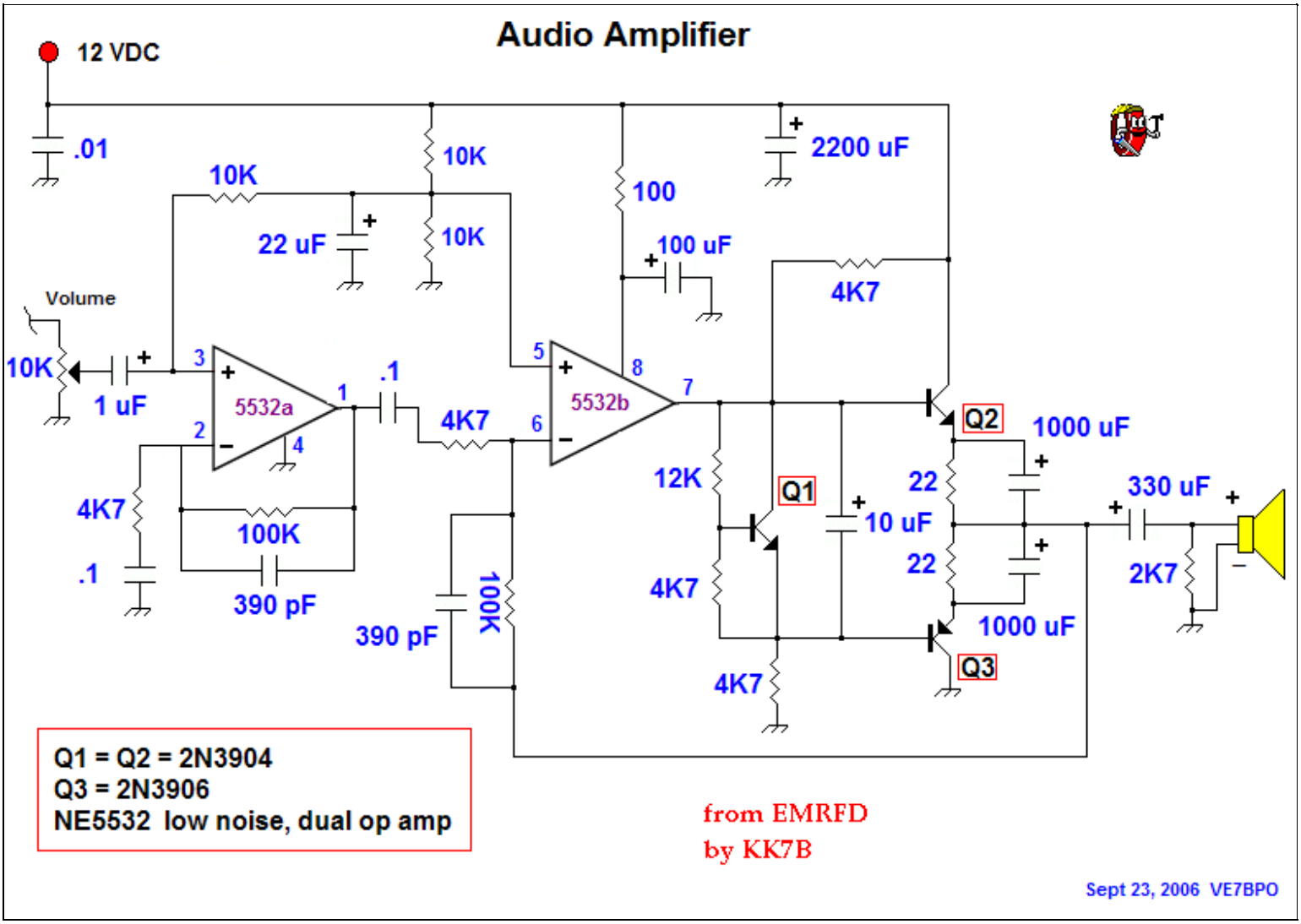
Felix called for a 1 mH radio frequency choke. The largest I had in stock was a 1000 uH choke. I had to decouple it as shown to prevent oscillations from occurring in my receiver. For the 1 uF and 2.2 uF capacitors, I used polyester film types which sounded better than electrolytic capacitors.

Oct 13, 2006: Note. The 1000 pf input cap to the detector was omitted in error in the original schematic which is now correct.



To the right: The detector board. On the left is an op amp preamp stage that was later disconnected as it was not needed. Note the copper is removed where the chassis mounting nut contacts the copper board. Both audio boards were isolated from chassis ground and star grounded to a single point. The speaker negative terminal was also directly connected to this point. There is no hum.

Audio Amplifier



Above schematic. Rick, KK7B designed this low noise audio amplifier. It is from EMRFD. This superb AF amp greatly compliments the VK4FUQ detector. This is the best speaker audio amp under 1 watt I have ever used. Distortion is very low as long as it is not over-driven. I increased some capacitor values compared to the original schematic. Please refer to EMRFD for details on this stage. The chassis of this receiver greatly increases the low frequency response. On the 1 second pulses of WWV, the receiver "knocks" like a metronome. This does not occur when the

chassis lid is off.

WWV web site: http://tf.nist.gov/stations/wwv.html All the often subtle pulses and tones transmitted at various times during the hour can be heard with this receiver.

To the right: A bread board of the AF amp. This is the audio amp I shall use in future projects which contain a speaker. Kudos to KK7B.

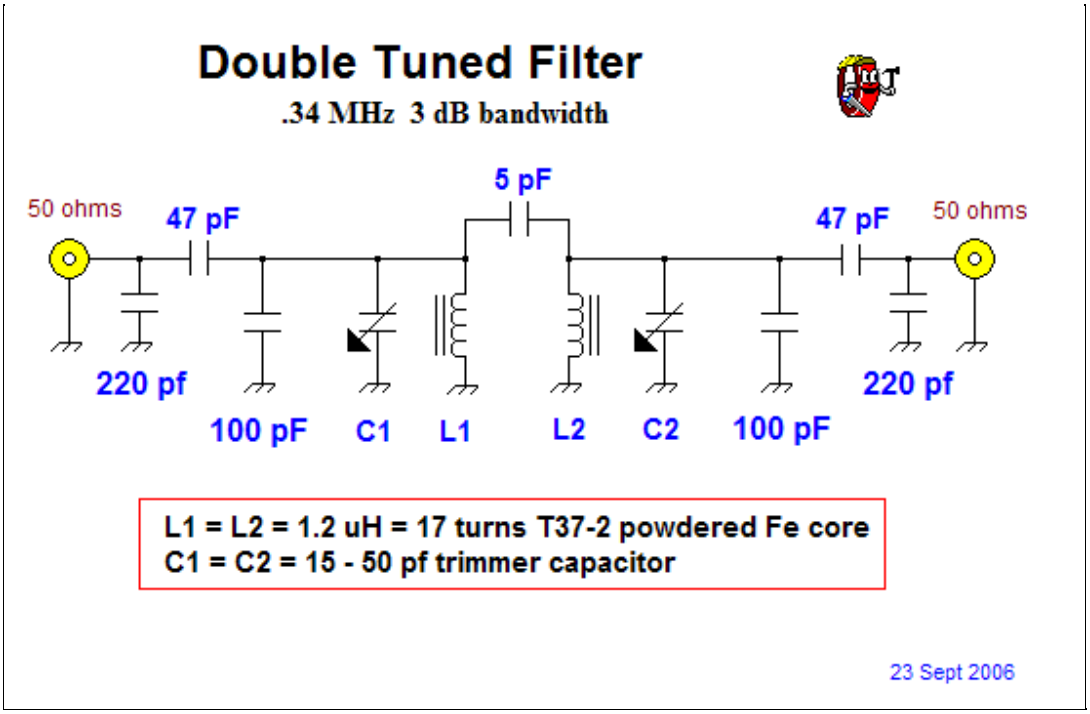
Below 3 images: Different views of the TRF receiver. On the front from left to right are the bias "sensitivity" control, volume control with integral power switch and blue LED "power on" indicator.

On the rear from left to right are the 12 VDC input jack, an unused switch (was an -10 dB attenuator at 1 point), the RF gain control and a coaxial SO239 connection.

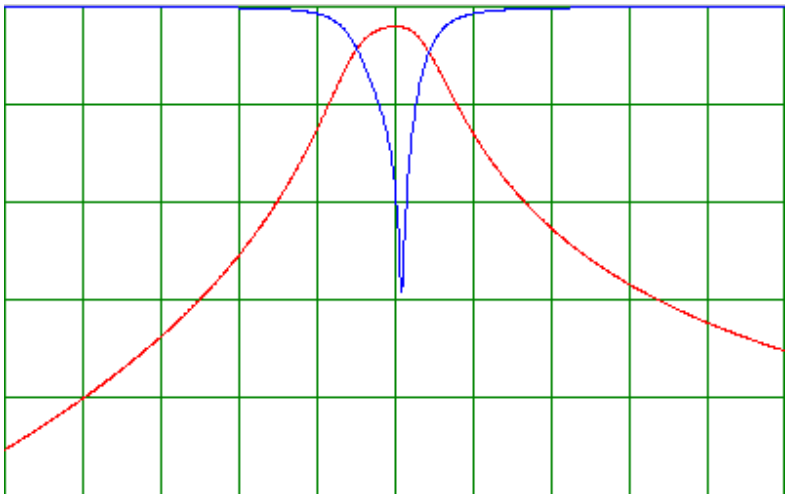
Further Experiments

What follows are some of the ideas and circuits tried over the past year.

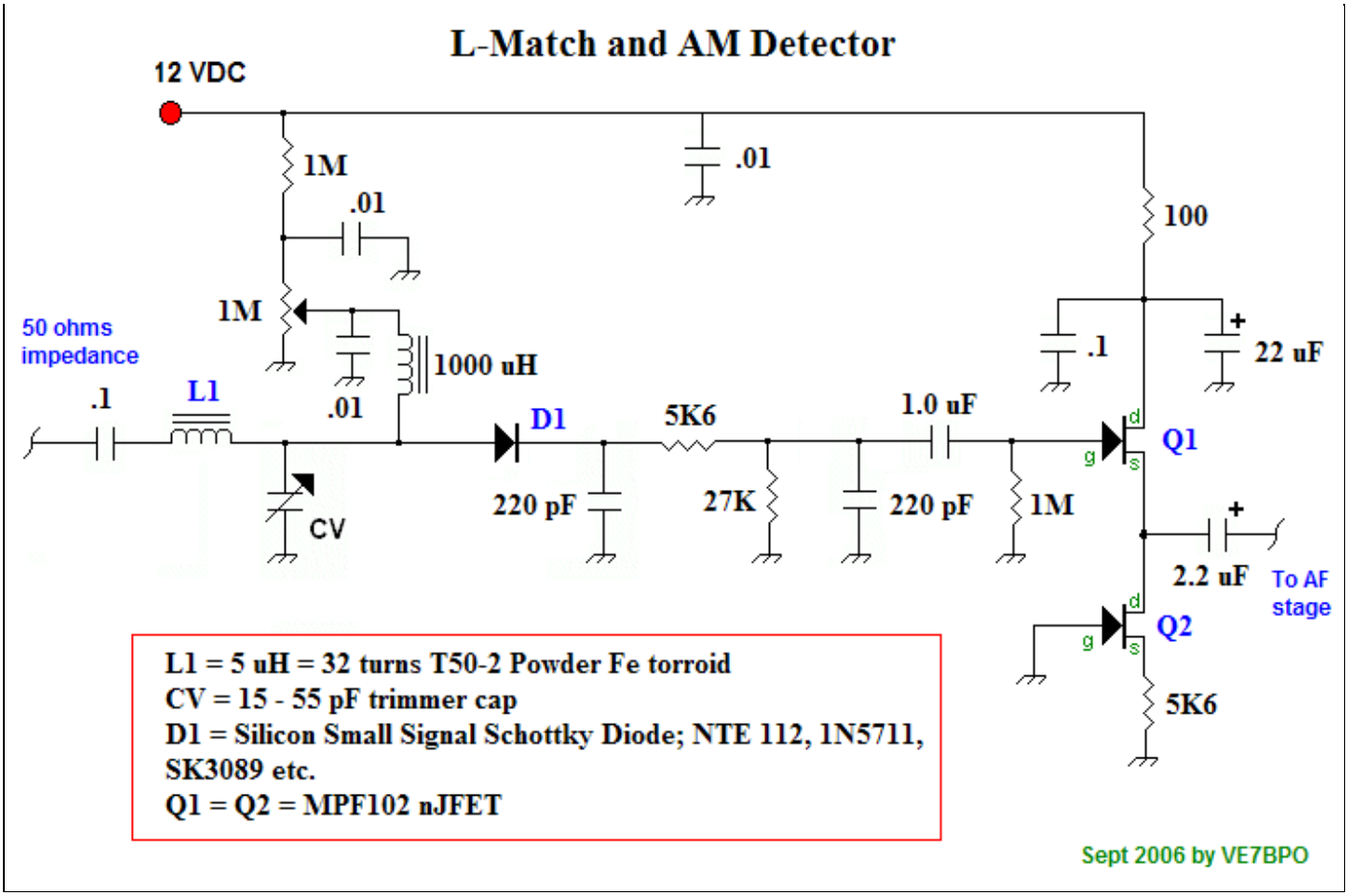
Schematic to the right: A 10 MHz, double tuned RF band pass filter that may be used ahead of the receiver. Insertion loss is ~ 3 dB and this filter uses a 5 pF coupling capacitor which are not too difficult to find. Filters with bandwidths of 150 - 180 KHz were also tested.



To the right: A GPLA simulation of the popcorn DTC shown above. titlehough a little mistuned, it is reasonable for a filter that uses common junk box values and has low insertion loss.



L-Match AM Detector

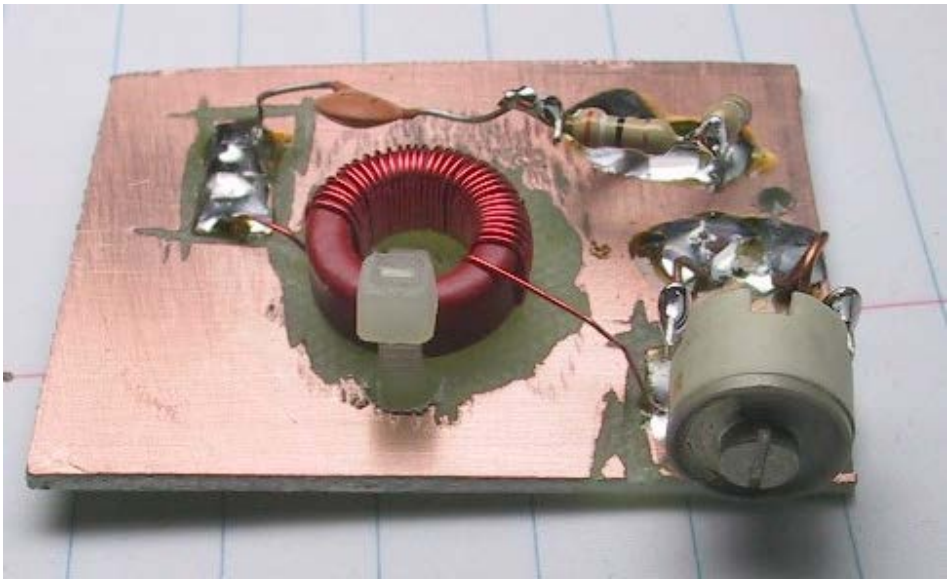


Above schematic. The VK4FUQ detector can also be used to follow a 50 ohm output impedance stage by using an L-match as shown. The L match tunes very sharply. I peaked the L-match with the bias at 0 volts.

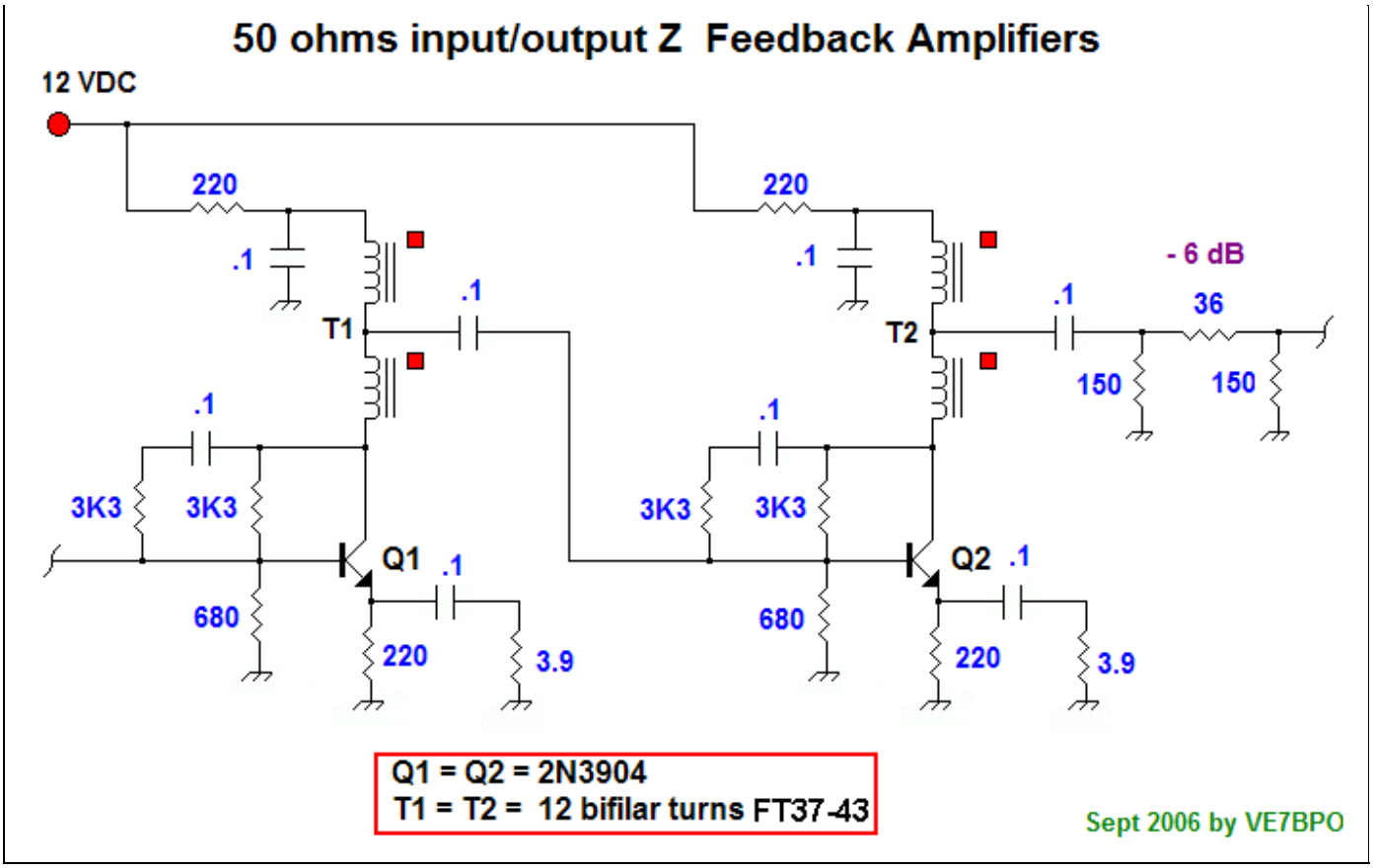
The input impedance of the detector is related to the DC current flowing in the diode. This is established by the adjustable bias current or "sensitivity control". The input resistance will be 26/I, where I is the current in mA.

For example, if the current in the diode is 10 microamps (0.01 mA) the input Z is 2600 Ohms. I have found that any input Z value from 2000 to 5100 ohms worked well with this detector.

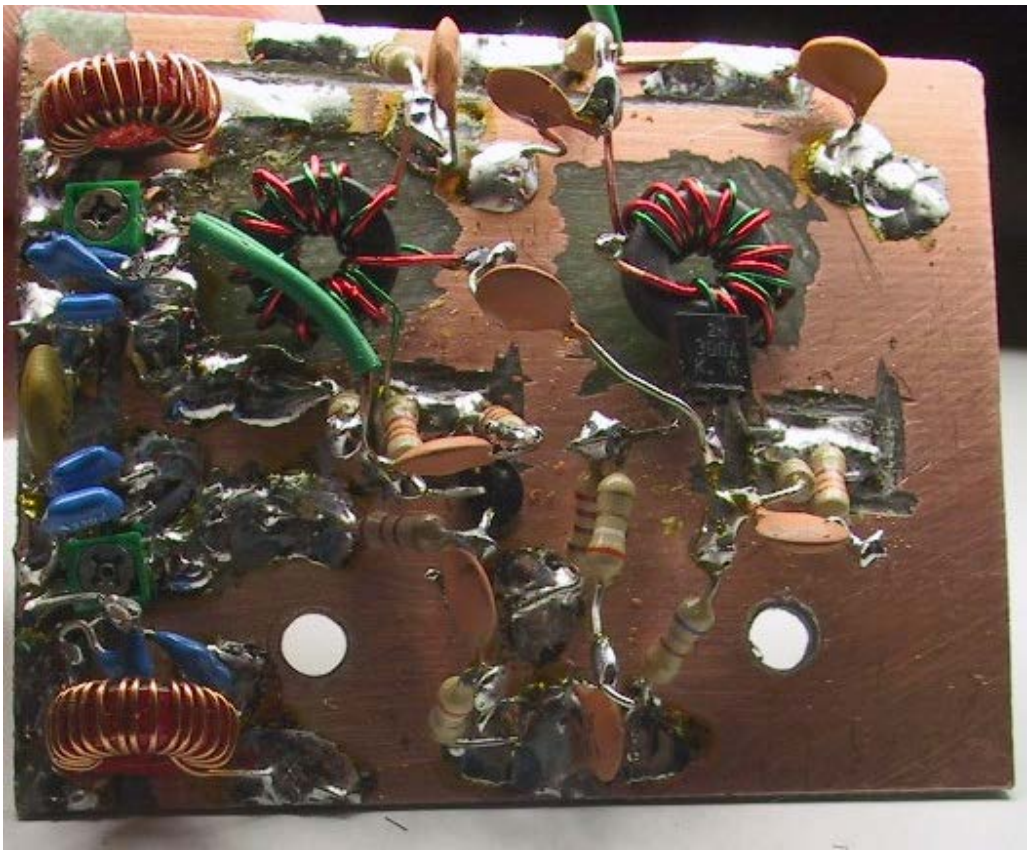
Image to the right: A photograph of the L-match connected to a - 6dB 50 ohm pad which terminated the 50 ohm feedback amplifier that drove the L-match.



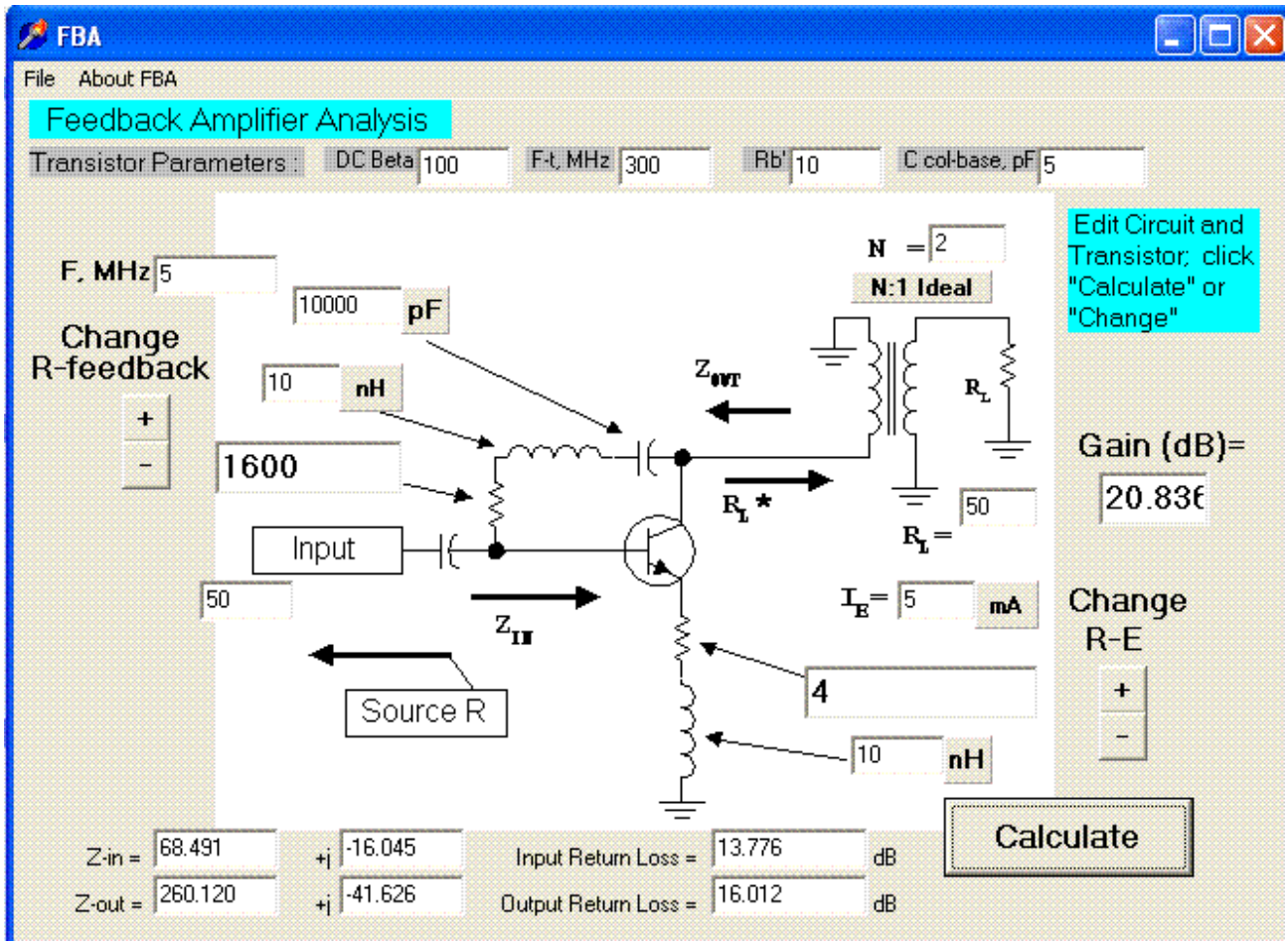
Feedback amplifiers



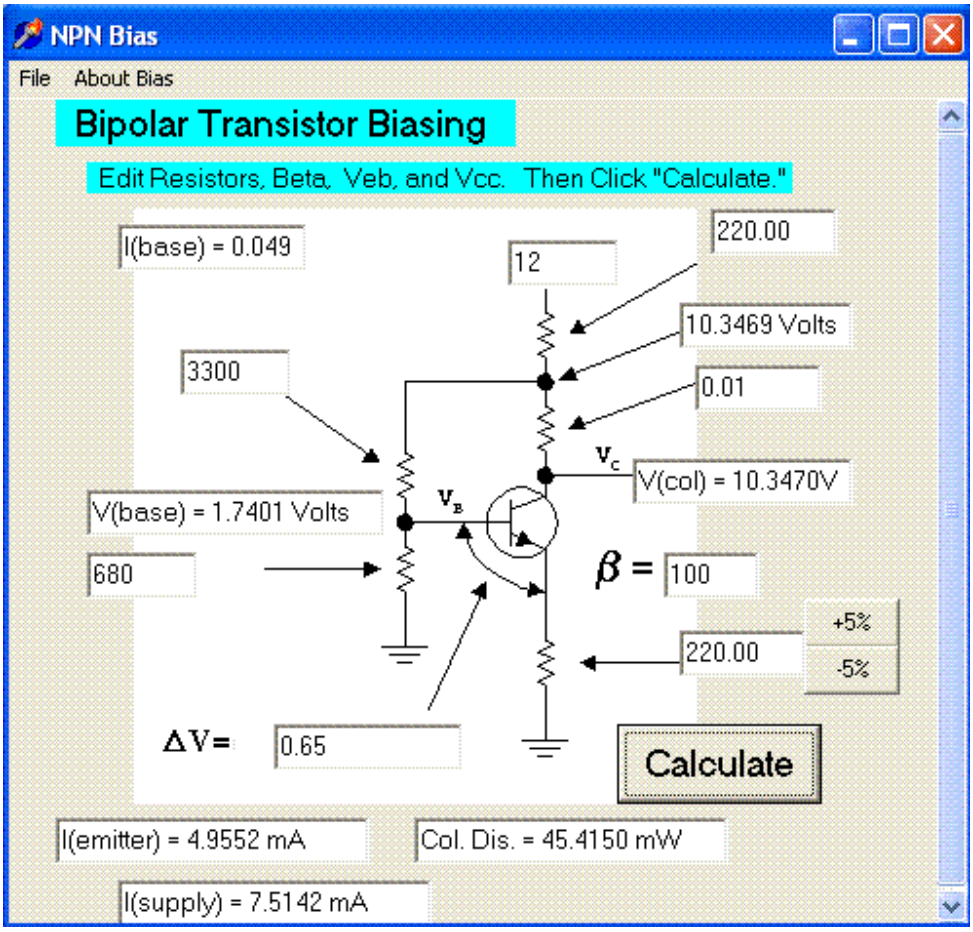
Above schematic. Feedback amplifiers may be used as RF amplifiers for a TRF receiver. This stage followed the crystal filter in one version of my TRF receiver. Stability was excellent. This feedback amplifier was designed by Wes, W7ZOI. It has ~ 20 dB gain and draws a little over 5 mA current.



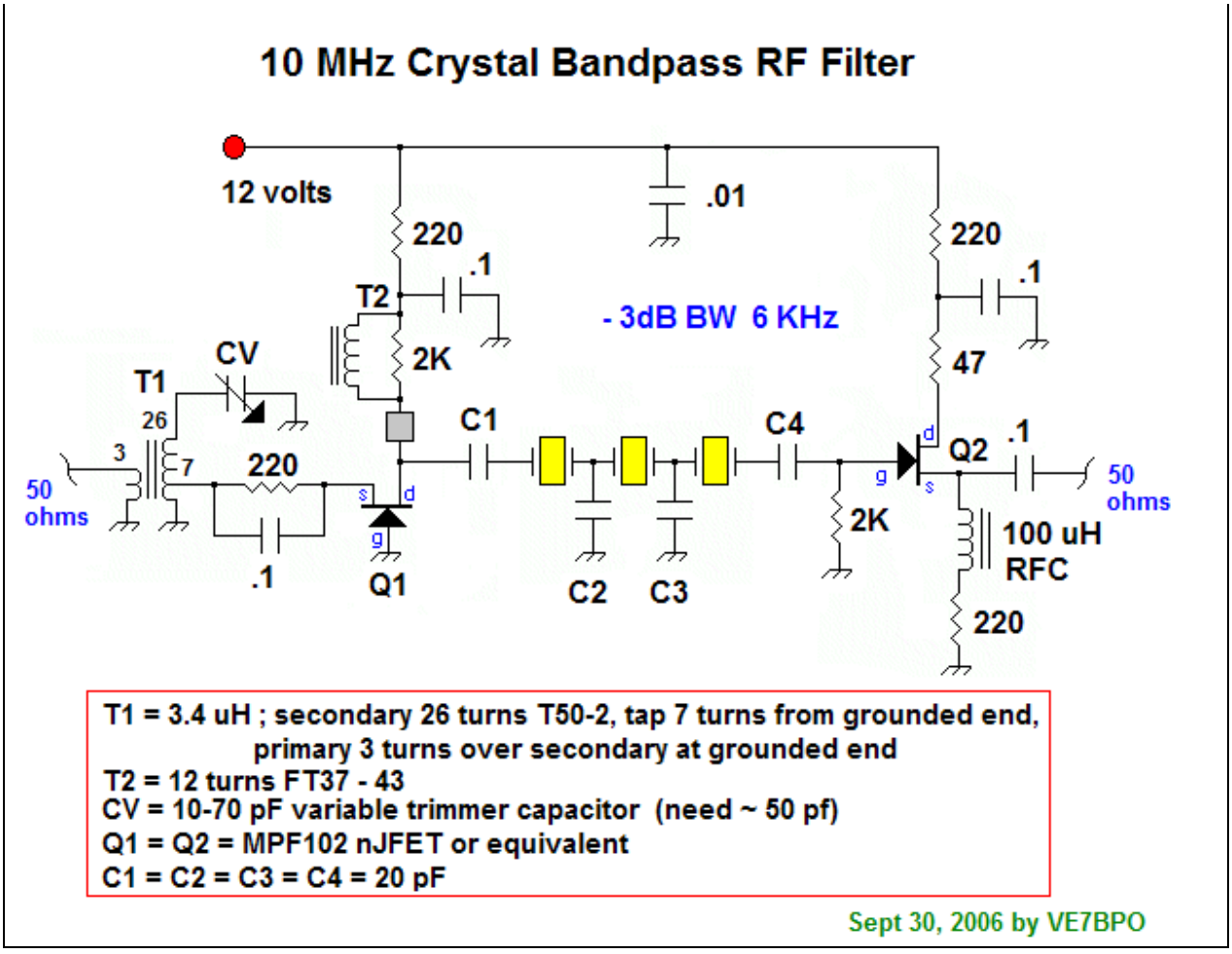
Above graphic. 2 feedback amplifiers are shown on this breadboard. The double tuned filter (DTC) shown earlier is also built on this board. In this version, the crystal filter was omitted and replaced with the DTC.



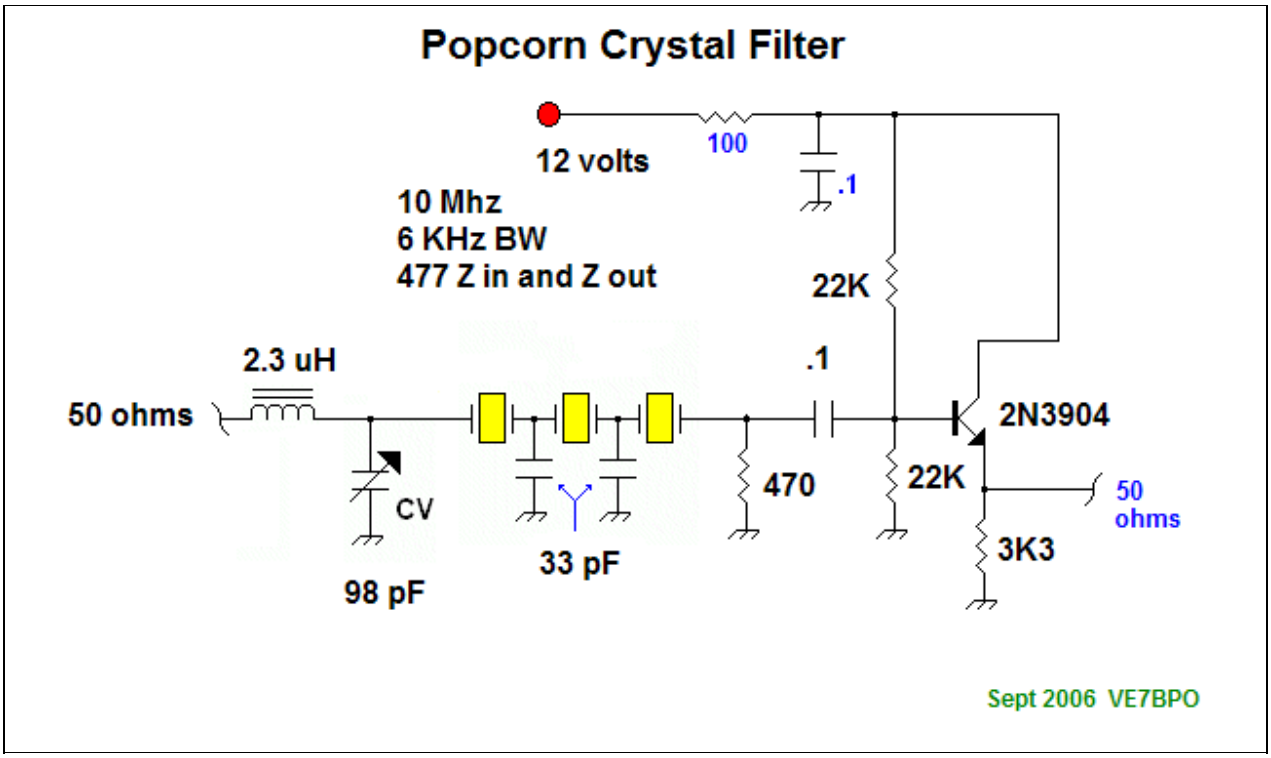
Above graphic. Using software that ships with EMRFD, W7ZOI designed the feedback amplifier used in this version of the receiver.



Above graphic. The feedback amplifier bias resistor values were also calculated using software written by W7ZOI and included with EMRFD.

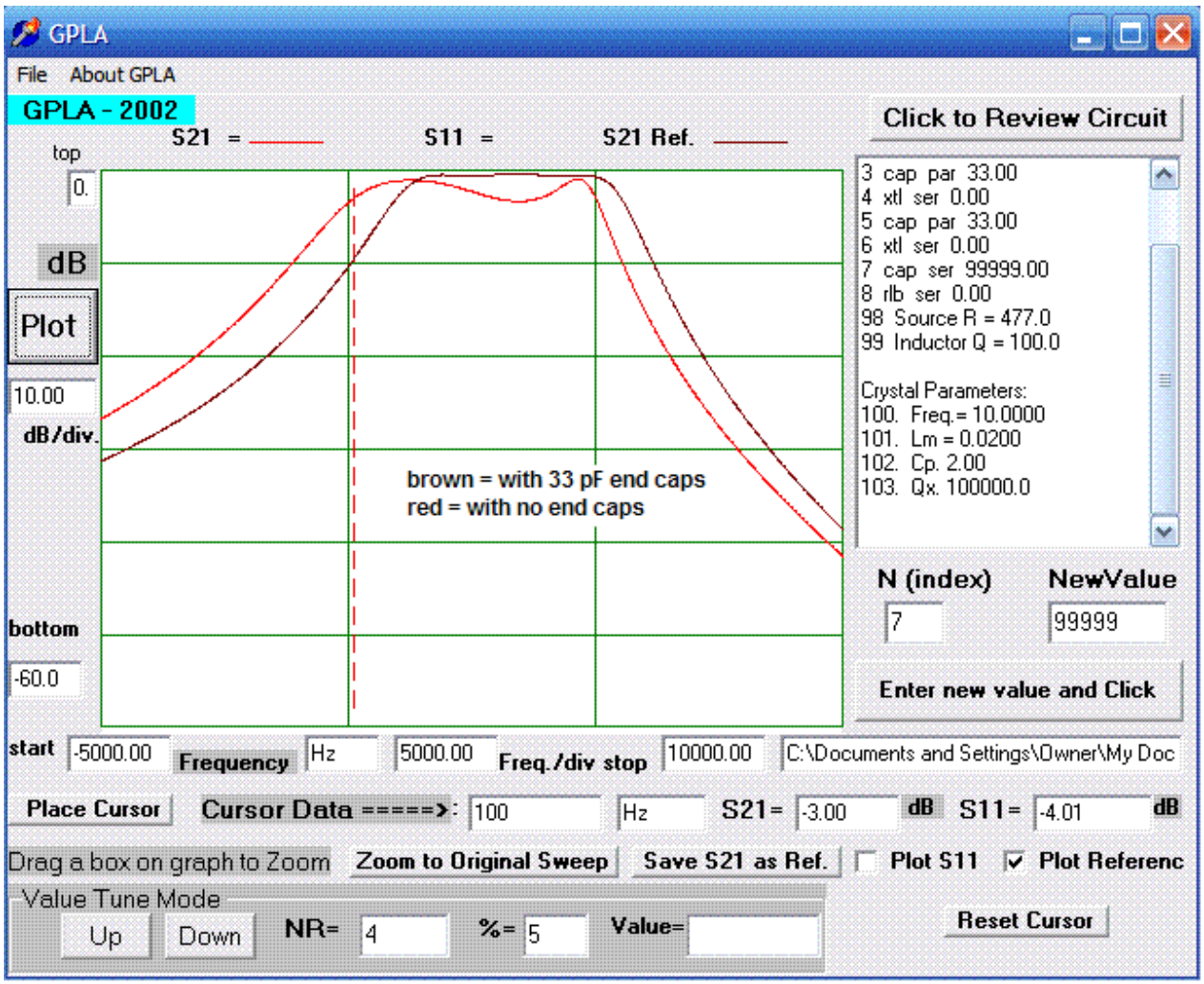


Above schematic. This circuit has the crystal filter matched to 50 ohms input and output stages using JFETs. The JFETs also serve to provide a little more gain. Careful layout is required to reduce BCB interference for all stages in a TRF receiver.



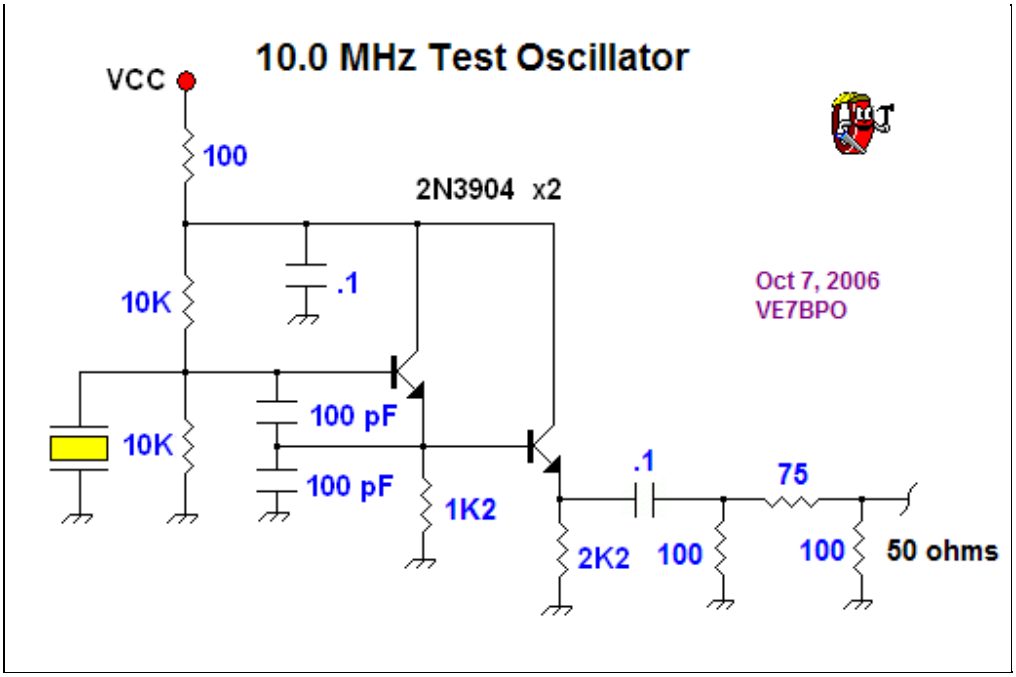
Above schematic. This is the original crystal filter that I designed. The input and output impedance is 477 ohms. The input was matched to the preceding 50 ohm stage using an L-network. A emitter follower is used to match the output to the 50 ohm stage which followed. Later, the emitter follower was replaced with the source follower (with a 470 ohm gate resistor) that is shown in the schematic directly above. The source follower

had greater immunity to BCB interference and provided a better termination for this filter. This popcorn filter worked well, titlehough occasionally there was another station in addition to WWV, in the pass band. This also happened with the filter used in the final version of this receiver.

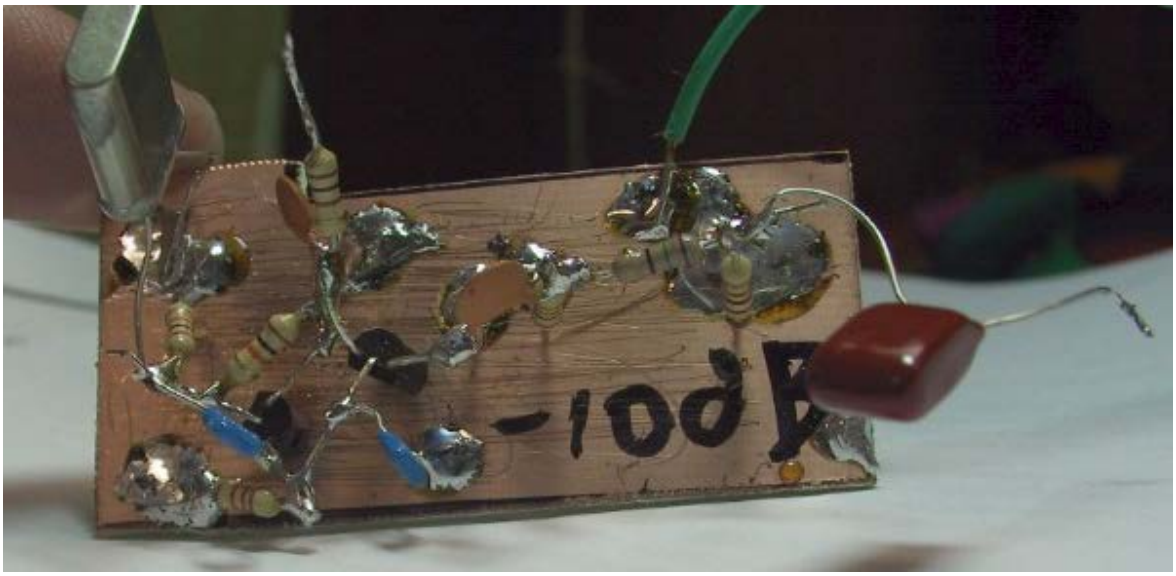


Above graphic. Here are GPLA simulations of the popcorn filter with and with out 33 pF series end capacitors which serve to tune the filter. The brown tracing illustrates that it is better to include a series 33 pF cap at each end of the "popcorn" crystal filter. I did not use this filter because the dual gate MOSFET RF amps used in the final version, have better gain driving or following the 2000 ohm filter designed by W7ZOI.

GPLA is a "must-have" program. You can "tune" filters with different or asymmetric input/output impedances.

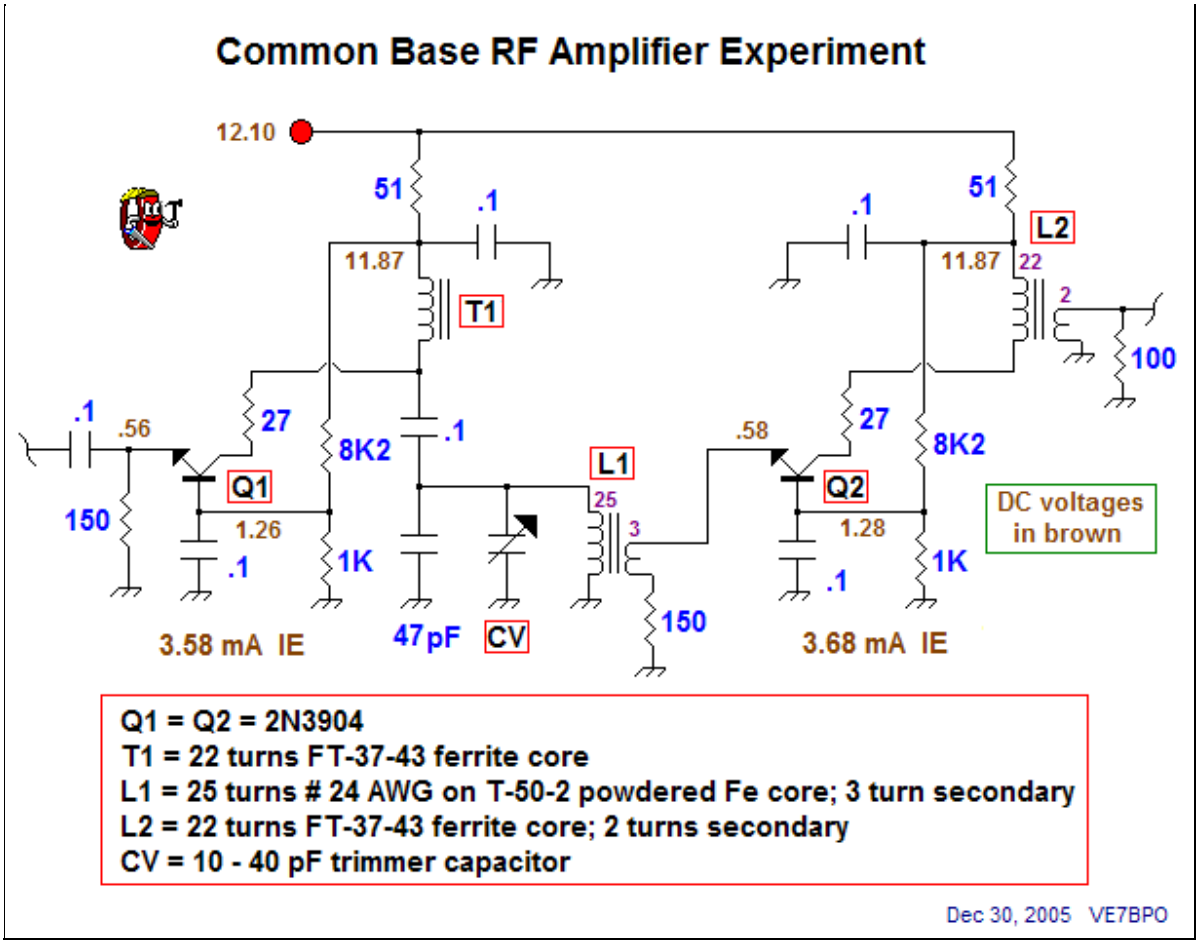


Above schematic. This 10.0 MHz crystal oscillator has a - 10 dB, 50 ohm pad on the output and was used to match the crystals, test the RF amps and align the filters used in these experiments.



Above graphic. A breadboard of the test oscillator shown above.

Above graphic. Some of the bread boards developed during experimentation. My final receiver layout (and potentiometer positioning) is not optimal, however this is a prototype and I had no idea what the finished version would look like.



Above schematic. This is one RF amp that was built for this receiver. The turns ratios on L1 is too drastic to afford much gain.

Above graphic. I have been told many times that my breadboards are very ugly looking. This breadboard of the schematic directly above, shows that occasionally, I can build a nice looking circuit!

Conclusion

The highlights of these experiments were VK4FUQ's detector and KK7B's AF amplifier.

When constructing such a receiver, *build backwards*. Install the speaker and then build and test the AF amp. Test it by touching your finger to the input and listening for noise or BCB radio. Turn the 10K pot and verify that the noise increases or decreases appropriately. Perhaps test it using an AF oscillator.

If it works, you get immediate positive feedback and motivation to continue. If it does not work, you only have 1 stage to trouble shoot.

Next, build the detector. To test it, connect a piece of wire about 25 cm long between the RFC and the anode of the diode. You should then hear local BCB radio. Slowly turn the bias potentiometer from 0 to fully on. Notice how increasing the bias may bring in 2 or more stations compared to when it was at 0. Also notice how it changes the tone and sensitivity of the detector. Try shortening the "test antenna" and observing how sensitive this detector is with the bias increased.

If all went well, you now have an AM radio!

Next add in the Q3 RF amp and again test it using a short piece of wire. Then continue on until you arrive at the antenna connection for your receiver.

Best regards, VE7BPO

MF and HF Receiving Antenna

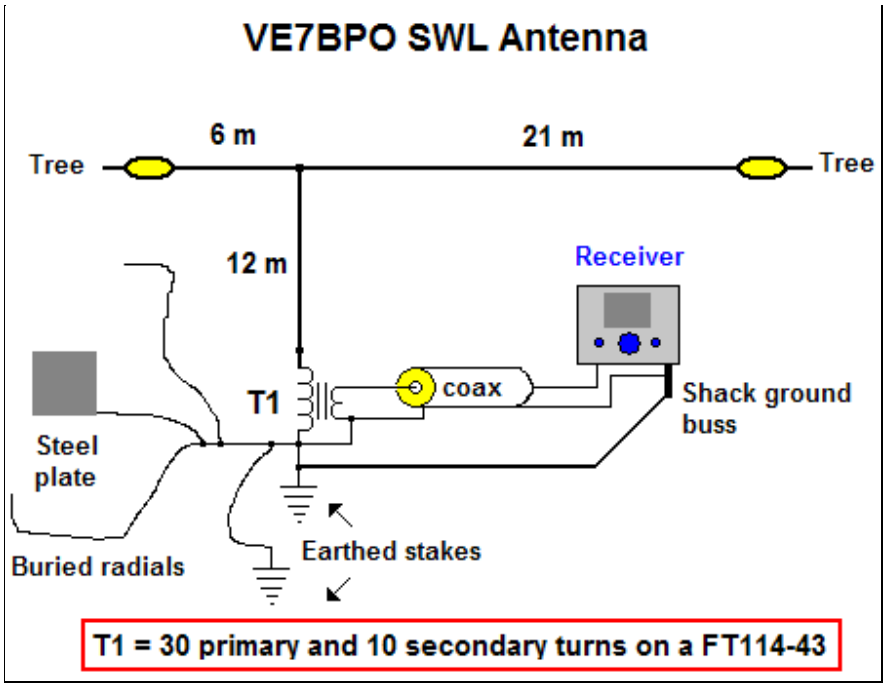
Introduction

My first shortwave antenna was a simple end-fed wire which started at my bedroom window and extended out horizontally to a tree which was 25 feet away from our house. The antenna feed line was a short piece of wire that connected to the near end of the antenna and entered the house through a small hole I made in my wooden window sill. This feed line was directly connected to my receiver's high impedance antenna input. My station ground was long piece of wire that was connected to a copper pipe located in the bathroom next door. While this antenna brought in "the world" to my bedroom, it was extremely noisy. Directly connecting your antenna feed line and house ground system to your receiver are not good RFI reduction practices. This web page will explore some experiments in trying to minimize the Radio Frequency Interference (RFI) arising from my local environment.

Indoor RFI sources are usually plentiful. Electrical appliances such as washing machines, televisions, DVD players, computers and electrical wiring may all emit RFI which your antenna, or directly connected house ground system may pick up and feed to your receiver. Certain indoor devices may be really strong RFI sources and will have to be eliminated or decoupled. Outside of your house are also potential sources of RFI. These may include such things as power transformers, electric fence and garage door openers. RFI location and reduction is out of scope for this web page, however a good place to learn more is the ARRL RFI book (out-of-print: search on Amazon). To find RFI sources in your home and neighborhood, try using a battery powered AM radio. At my QTH, I located a noisy VCR inside the house my Grundig S350. We rarely use this VCR and now just leave it unplugged until we actually need to operate it. I tuned the receiver to an empty frequency and found this VCR by trial and error. **Please note this web page is concerned with feeding a shortwave listening antenna and does not describe providing protection against lightning.** For web sites which covers lightening plus RF ground please refer to this offering from W8JI or eHam.net.
Protect your home and family from lightening !!

Outdoor MF and HF Antenna

The schematic to the left summarizes the outdoor VE7BPO MF and HF receiving antenna system for summer 2007. Although modest for a big city lot, this antenna seems to pull in the DX and is relatively free of RFI. This antenna was just a case of "putting as much wire in the sky as possible" and the dimensions are indicated for interest sake only. The 27 meter long horizontal section is supported between 2 trees at a height of about 14 meters high. The weight of the



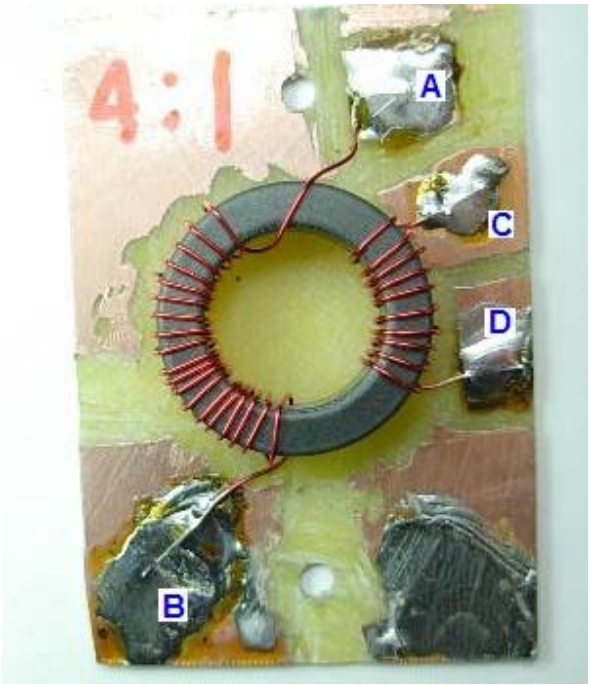
vertical element wire plus slack in the horizontal wire droop it to about 13 meters high in the center. The vertical section is soldered to the horizontal wire 6 meters from the nearest anchoring tree and runs straight down to the antenna feed point which is about 1 meter off the ground. The feed point is a piece of copper-clad PC board (with isolated sections created with a hobbyist motor tool) and is bolted to a long copper pipe which serves as the first station earth-grounding stake. A transformer (T1) configured as a UNUN **(unbalanced-to-unbalanced)** is used to interface the antenna with 50 ohm coax that runs through the house and into the radio shack. Some rudimentary experiments with the UNUN and the earth-grounding system were undertaken.

The methods I used to potentially lower unwanted RFI to my antenna system are as follows:

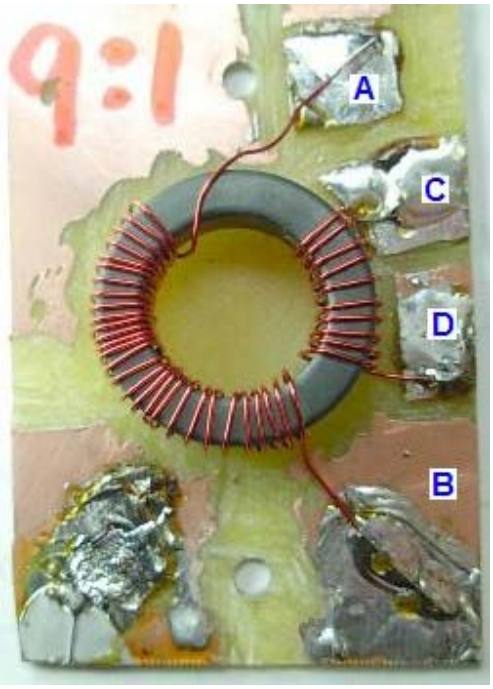
1. The receiver and power supply are independently connected to a single, central ground point (ground buss) in the radio shack.
2. 6-10 gauge wire is used for my ground system (not including the radials which are bare 12 gauge wire).
3. The ground wire connecting to my first earth stake to the station ground buss is just outside the shack window and is short as possible to provide a low impedance and low inductance path for MF and HF frequencies.
4. There is a second ground stake located 1 meter from the primary ground stake (I will add 2-4 more in time).
5. I have a large piece of steel buried underneath the soil tied in to my system as well as 3 bare copper radials. The radials are 3 - 7 meters in length.
6. New RG58/U coax was used as the feed line.
7. All wire splices in the grounding system are soldered and taped up. I used conductive grease (to prevent oxidation at the wire-stake interface) on any clamps connected to ground stakes. My ground stakes are ~ 2 meters long.
8. The earth grounding area soil is moist and peat-laden and is watered regularly.
9. I plan to maintain this ground system every 2 years.

4:1 UNUN

My antenna is almost an end-fed wire with both a vertical and horizontal section. I do not have the gear to measure the impedance versus frequency in the MF and HF bands. I do know that on some bands it may present an impedance of several thousand ohms and a transformer can smooth out the variation in impedance versus frequency so my receiver sees a relatively low impedance on most bands. The transformer also serves to help reduce RFI from my antenna system by eliminating unwanted common mode currents flowing on the outside of the coax braid. Grounding the antenna via the UNUN will also prevent static electricity from building up on the antenna. My first UNUN had a 4:1 impedance ratio. It is shown to the right. The antenna connects to point **A**. The ground stake connects to point **B.** Point **C** connects to the inner wire of the coax and point **D** is connected to the braid and also the grounding stake. I used 24 AWG wire and A FT114-43 ferrite core. You can clearly see there are 20 primary windings and 10 secondary windings loosely coupled. I chose the FT114-43 core because I had it on hand and the 24 gauge wire provides good mechanical support for the coil. I could have used an FT50-43 ferrite torroid as well with a smaller wire gauge. You can also use a bifilar transmission line type transformer. I was very happy with this UNUN and however it did not have as much signal strength as I expected on the 160 meter amateur band and below.



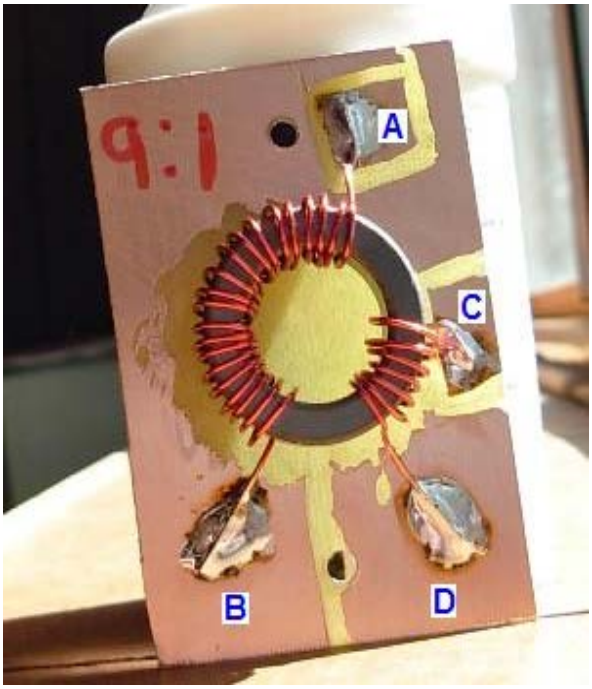
9:1 UNUN



Next I tried a 9:1 impedance ratio UNUN. This an extremely popular impedance transformation ratio for end-fed or random wire SWL antennas. I wound 30 primary and 10 secondary turns on a FT114-43 ferrite torroid. The connection points are identical to those described in the above 4:1 UNUN. Remember that the impedance transformation ratio is the square of the actual turns ratio on your transformer; thus my 3:1 turns ratio is a 9:1 impedance ratio. Electrical engineers commonly use a rule when winding broadband transformers such as these. The inductive reactance (XL) of the smaller winding must be at least 4 times the load impedance at the lowest frequency that the transformer "looks" into. So for 50 ohm coax, the XL should be at least 200 ohms at 500 KHz which is the lowest frequency I intend to receive. The formula for XL is **XL = 6.28 X F X L**. Frequency (F) is in Hertz and L is the inductance in Henries.

At 500 KHz my inductor has an XL of 189 ohms which is almost perfect. I should have used 11 turns which is an XL of 229 ohms and strictly observes the 4X rule. Therefore my UNUN ideally should have used 33:11 turns on the FT114-43 torroid. If you use a FT50-43 torroid, use the same 33:11 turns ratio; this will provide 198 ohms XL at 500 KHz. For practical purposes, my 30:10 UNUN should work fine as I rarely tune frequencies less than 1000 KHz. I found this UNUN to have strong signals all the way down to MF and decided to use a 9:1 impedance ratio for my antenna system. Many experimenters and a few commercial UNUNs recommend the 9:1 impedance ratio for multiband end-fed or random wire antennas. Eventually I will encase it in a water and UV proof enclosure.

Conclusion



My experiments while constructing a reasonable quality MF and HF receiving antenna confirmed that using a UNUN, coax and a good RF ground system can reduce common mode RFI in my receivers.
I also tried temporarily connecting my ground system to a copper water pipe located in my shack while listening to WWV at 5 MHz and immediately the noise level rose 2 S-units on my receiver.
This pipe was clearly not grounded in my house where there is a mixture of plastic and copper water pipes. Additionally, my antenna wire and feed point is away from the house in a quiet area according to listening tests using a Grundig S350.

It is relatively easy to construct a UNUN on your bench using a ferrite torroid. Many builders have emailed me to say they do not feel comfortable winding torroids. Torroids are easy to use and by winding a couple and experiencing some success, your confidence working with them will surely improve. If you live in a part of the world where you can not easily obtain a suitable ferrite core, just email me and I may send you an FT50-43. You can also choose a ferrite with a different core permeability. Some builders use number 75 material. I used the FT114-43 because I get all my torroids from W8DIZ and just use what he has in stock for my projects. If you really do not want to construct a UNUN, commercial products are available on the web on sites such as http://www.arraysolutions.com/Products/baluns.htm.

I wish you good luck with your own antenna experiments and please be safe!

Some SWL Antenna Related Links

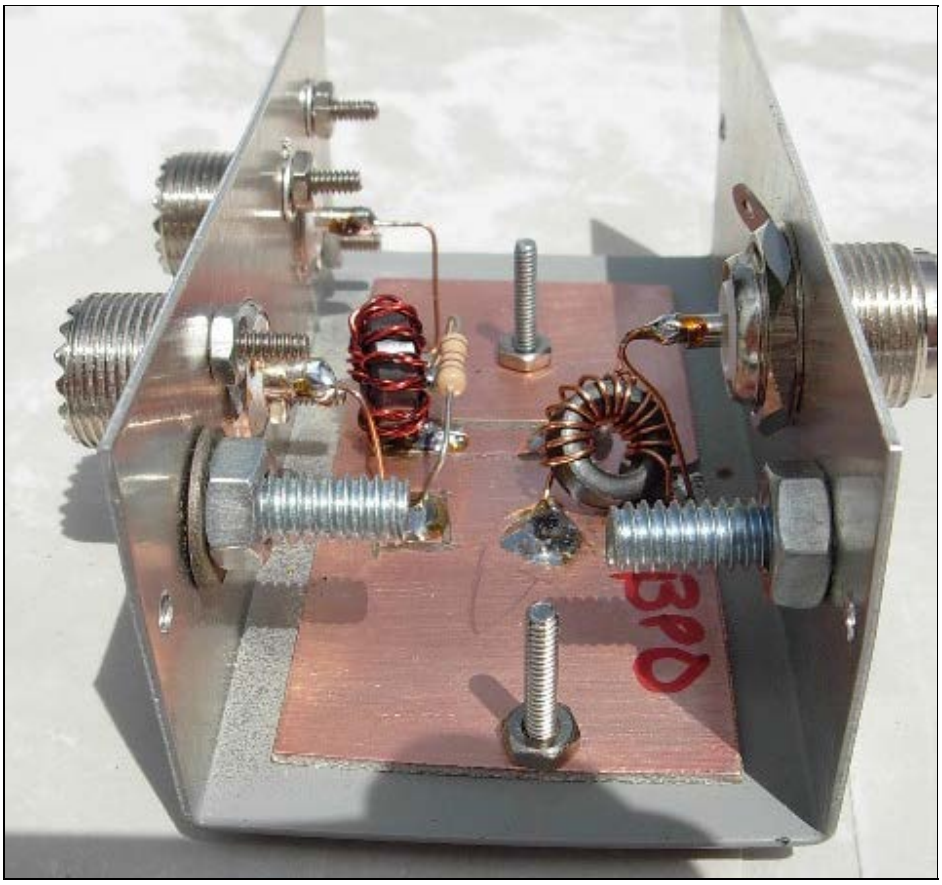
L.B. Cebik, W4RNL was a respected antenna expert. There is great information on his web pages

Build a Shortwave Antenna. A good overview of home brew multi-band antennas by N4UJW

MF and HF Receive Antenna Splitter

Introduction

As a radio experimenter, I have numerous MF and HF receivers to listen to but usually only 1 main outdoor antenna. Typically, this means that only my main radio receiver is connected to the outdoor antenna and my other receivers must use small indoor antennas with or without RF preamplifiers. I wanted to to permanently connect my main radio shack receiver and the receiver in the room directly above the shack to my main MF and HF antenna at the same time. The solution was to build a simple antenna splitter which allows the 2 radio receivers to connect to the single coaxial antenna feed line while preserving the correct impedance at all connection points. This project is based upon the splitter presented in EMRFD labeled Figure 3.81.

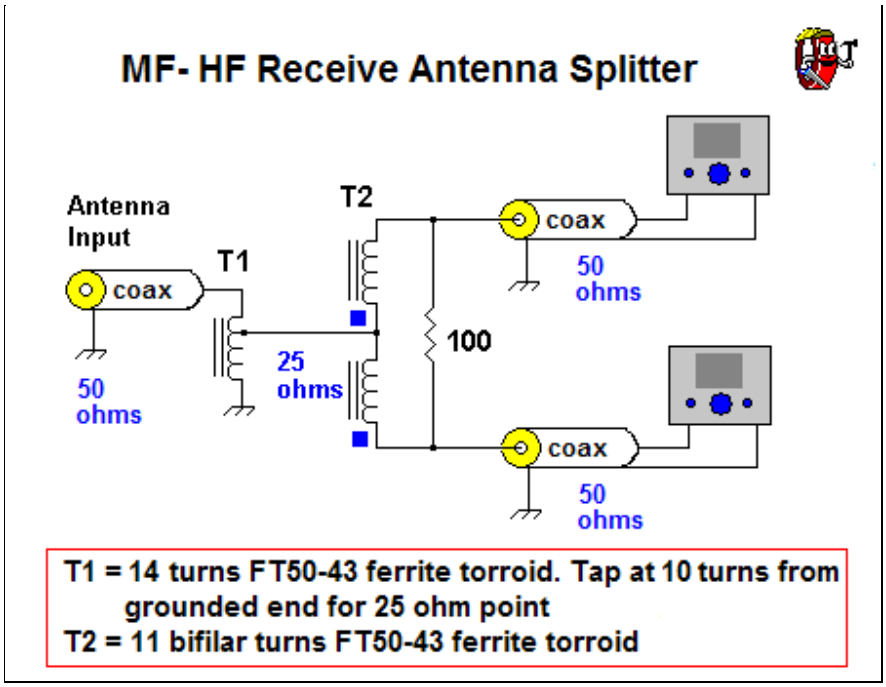


Each receiver and the antenna feed line have a 50 ohm characteristic impedance. This in-phase splitter is passive and has a loss of just over 3 dB. It is designed to operate from 500 KHz up to 30 MHz.

Please do not transmit through this device.

Project Schematic

The schematic to the left illustrates the entire splitter network from the antenna input to the input of the 2 receivers. T1 and T2 are broadband ferrite transformers with enough inductive reactance to tune down to the bottom of the broadcast band. If you only require a splitter for HF, then wind T1 with 10 total turns and a tap at 7 turns from the grounded end and T2 with just 10 bifilar windings. I used FT50-43 cores to allow the use of



thicker gauge wire which provides reasonable securement of the coils without external anchors, and because bigger inductors are easier to photograph. The FT37-43 ferrite core would also be a good choice, especially if miniaturization is a design goal.

Circuit Building Details

The antenna splitter breadboard is shown to the right. I used 3 colors of 22 gauge enamel covered wire to make my inductors. T2 is the actual splitter network coil and is the lower transformer in the photograph. The 100 ohm resistor serves to isolate the 2 ports connected to the receivers and absorbs impedance mismatches which may present when one terminal is not properly terminated. Note that the characteristic impedance at the input of this 3 port network is only 25 ohms. You can choose to ignore this or use an additional network such as a broadband transformer or an L- match between your antenna coax and the splitter to match this 25 ohm impedance.

I chose to use T1 which is an autotransformer with a tap at approximately the 25 ohm point. The splitter network worked well during my tests. Having only 1 receiver versus 2 receivers connected made no difference to the signal strength due to the excellent output port isolation.

Note that Wes, W7ZOI uses this 3 port network network several times in EMRFD. One example is the Lichen transceiver while another is the 6M superhet receiver presented in Chapter 6. Consult EMRFD for further discussion of this and other multiple port networks.

T2 is a bifilar transformer. The 2 wires were twisted together by securing one end of the 2 wires in a vise and the other end of these 2 wires in the chuck of a brace and bit (manual) hand drill. I twisted the hand drill until I had 8 twists per inch on the 2 wires . I used 2 color wire for ease of construction, however, it is almost as easy to tell the windings from one another by using an ohm meter or audible continuity tester.

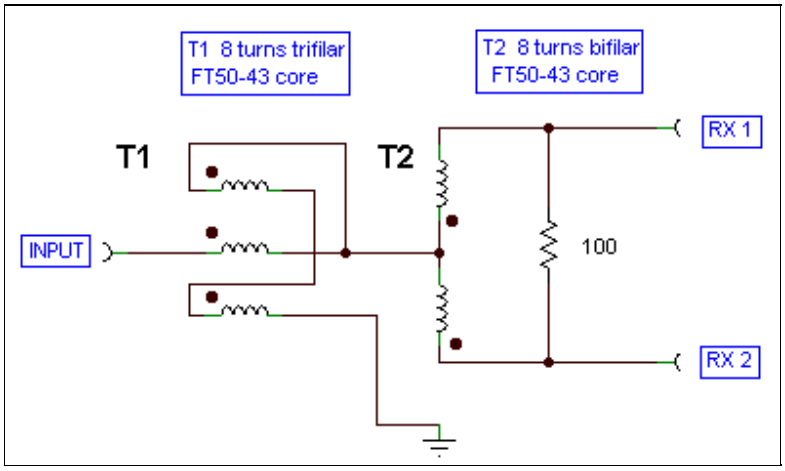
Chassis

To the right is the completed project showing the SO-239 connectors which are wired to the antenna splitter output ports. A chassis from an old project was recycled for this new project. The large bolts seen in some of the of the photos were used to fill in holes which had been drilled for the old project. This was done to provide improved RF shielding. The bolts also increased the weight of the chassis and help keep it from tipping over. Although it does not look as attractive as if I had used a brand new project chassis, considerable cost savings were realized. These little Hammond project boxes are getting very expensive. Also the

splitter is kept an the back of my main radio desk where it is out of sight anyway.

Update August 10, 2008 - Contribution by Dave, G4AON

This original network was designed for use in the MF to HF spectrum. Limiting this network is the input matching transformer T1 which negatively effects the T2 output port isolation; especially at 41 meters and higher. Testing by Dave, G4AON confirmed this. Dave designed, built and tested a trifilar wound, UNUN input matching transformer which provides a much flatter response for T2 port isolation from 0.1 to 52 MHz.



To the left you can see the G4AON input circuit for T1. In keeping with a design optimized for higher frequencies, less total turns are used on the transformers. His trifilar wound input transformer version is going to generate an impedance of (16/24)^2 x 50 = 22 Ohms at mid-band. My variation will generate (10/14)^2 * 50 = 25.5 ohms at mid-band. Using his version of T1 as opposed to my simple auto-transformer, Dave was able to provide better isolation of the output ports than the original design across a wider range of frequencies.

In the two popular, commercially sold RF splitters we have examined, the company did not even bother to match the input to the T2 transformer and some builders have written me to say they just omitted T1 and for their typical SWL listening this worked out fine for them. Increasing isolation across a wider frequency band and also matching the T2 input are issues that you the builder will have to consider. Certainly the lossy and often non-predictable #43 ferrite material is a factor which might affect your transformer performance. While a trifilar transformer is a little more difficult build for a novice as compared to an auto-transformer, this improved design might work very well at your QTH. Testing like Dave did is certainly the way to go and I greatly appreciate his contribution.

To the right are Dave's excellent bench measurements. He used a Marconi 2018 signal generator, a Racal 9301 RF millivoltmeter and a Bird load on the other port. Kudos to Dave for performing this experiment and contributing to the receive antenna splitter knowledge base.

Dave's web site.

Antenna splitter measurements		
Freq (MHz)	Port 1 & 2 loss(dB)	Port 1 – 2 isolation (dB)
0.1	-3	16
1	-3	31
2	-3	33
3.6	-3	32
4	-3	32
5	-3	30
6	-3	28
7	-3	28
10	-3	25
14	-3	23
18	-3	22
21	-3.5	21
24	-3.5	20
28	-3.5	19
52	-4	16

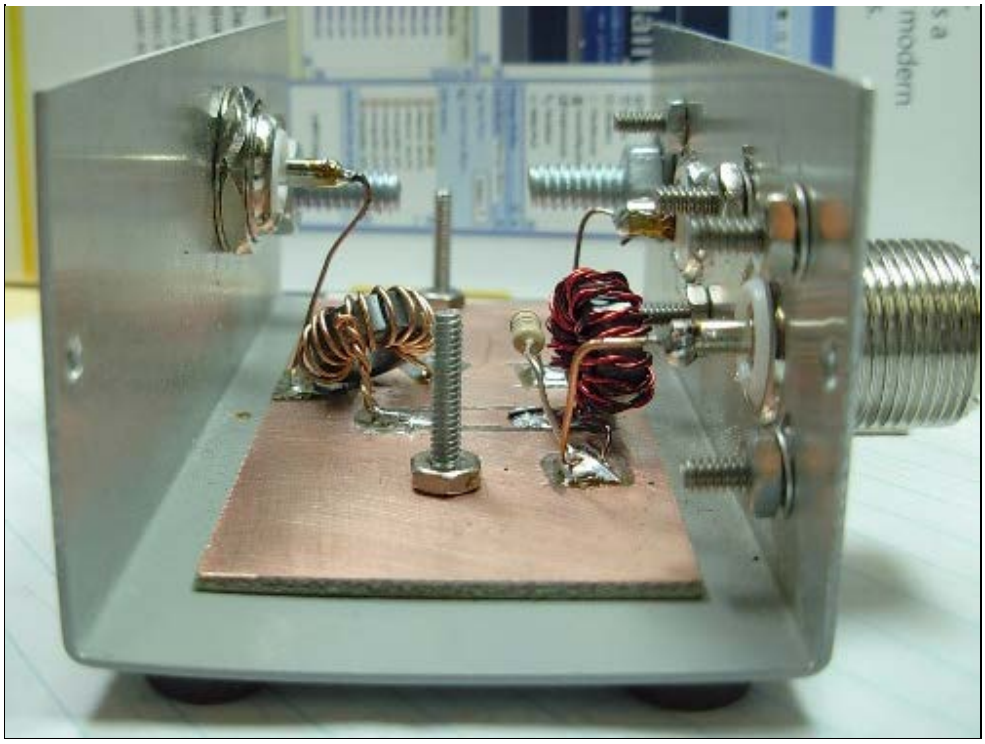
Conclusion

It is really awesome to be able to connect 2 receivers to the same outdoor antenna. The 3 port network and cabling to the additional receiver does not seem to increase receive noise levels from RFI in the house. Most likely this is due to the fact I am using shielded coax, a shielded project box and have a good RF ground system.

This is a simple project you can build in one evening.
I hope you receive some good DX! 73 es CUL, VE7BPO

Here is a link from F6AOJ

Additional Photos

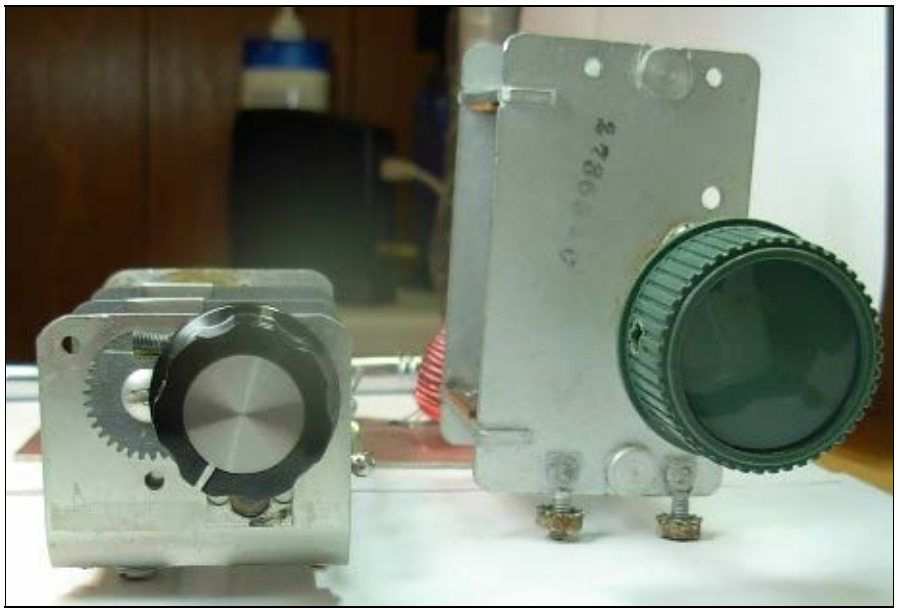


Medium Frequency TRF Receiver

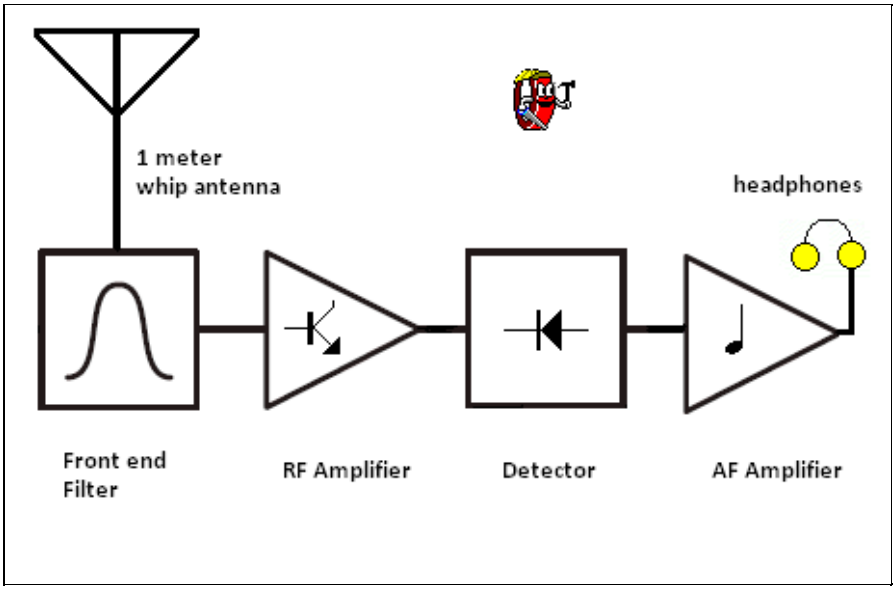
Introduction

This series of experiments was initiated in 2006, stalled, and was finally completed 16 months later with the inspiration provided by work regarding zero power receivers web-published by Wes, W7ZOI in late summer 2007.

Described is a complete receiver, built and presented backwards from the audio stage to the antenna. The design goals were to build a Tuned Radio Frequency broadcast band receiver with one RF amplifier, a high performance detector and a simple, headphone-level audio stage.



Receiver Block Diagram

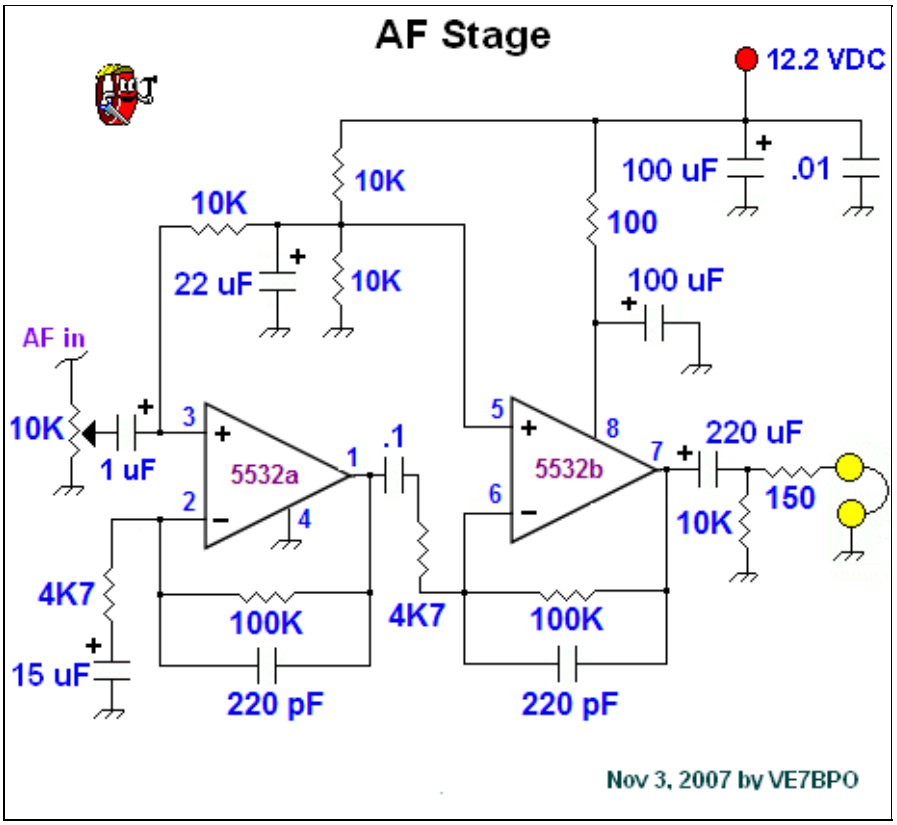


The receiver block diagram is shown to the left. The antenna is a ~ 1 meter long whip purchased from Radio Shack in the USA. A single cascode bipolar junction transistor amplifier boosts the RF voltage and drives an envelope detector which is terminated by a JFET source follower. The source follower connects to a 10K volume potentiometer which controls the AF signal voltage into a headphone-level audio amplifier. Like most of the projects on this site, the intention is to present some circuits and ideas for experimentation. This receiver is designed for local broadcast band AM radio reception, however, the various circuits could be used in or titleered for DX receivers as well.

A Supplemental Page can be found here

Audio Stage

The AF amplifier is a superb design by Rick, KK7B and is featured in many projects in EMRFD. This audio amplifier uses one 5532 op amp and has low noise and high gain. The 220 pF feedback capacitors can be increased to boost the low frequency response. I have built 6 or 7 versions of this stage and have used feedback capacitors up to 560 pF for this purpose. In the audio path, polyester film capacitors were used to try to improve the audio quality. Additionally, the value of the 15 uF capacitor connected to pin 2 is flexible. The quiescent current draw of this stage at 12.2 VDC is 12. 3 mA. Some builders may have to increase the 100 uF filter capacitor on the main 12.2 volt line to overcome motorboat oscillation. None occurred in my breadboard version. I suggest using this audio stage instead of the LM386 or discrete component final audio amplifiers in all projects which call for a headphone-level audio power amplifier on this web site.



Close up of the KK7B audio amplifier breadboard from the 10K potentiometer to headphone jack

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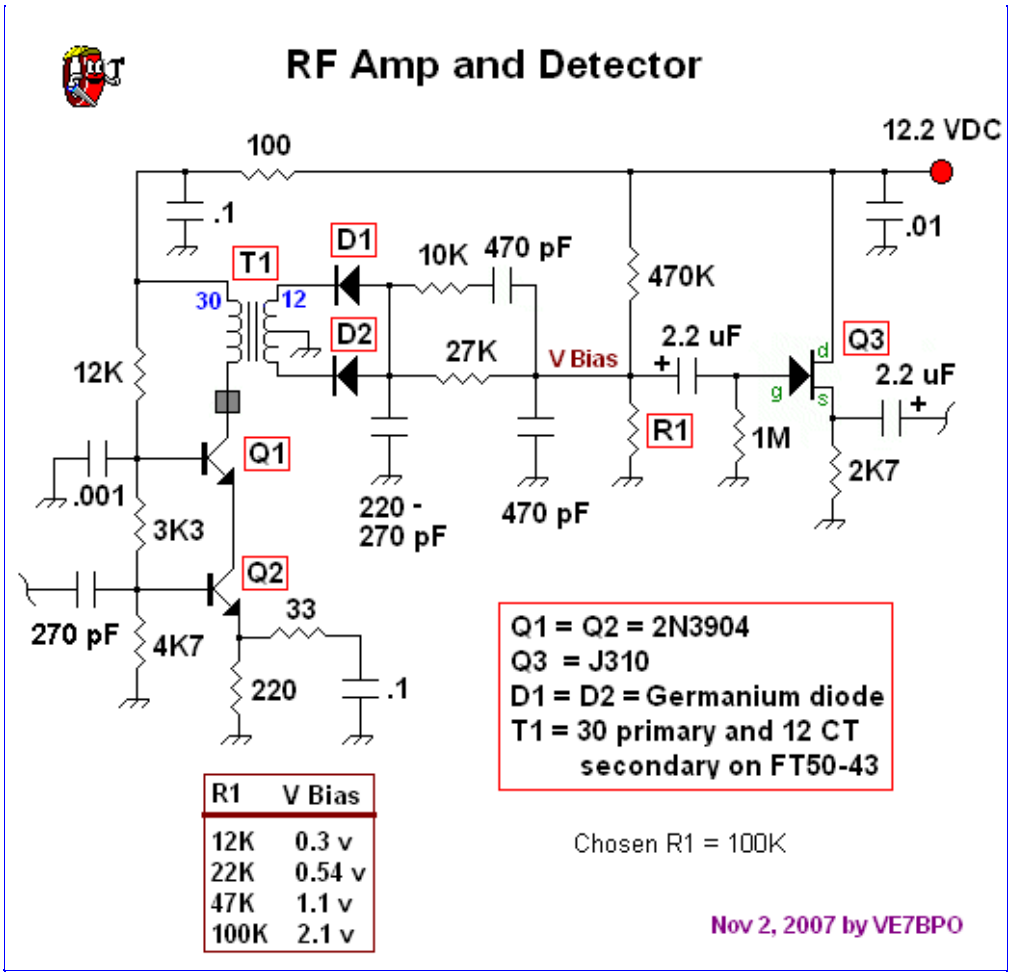
Cascode BJT RF Amplifier and High Performance Detector



Above is the combined RF amplifier, detector and JFET source follower schematic.

Cascode BJT RF Amplifier

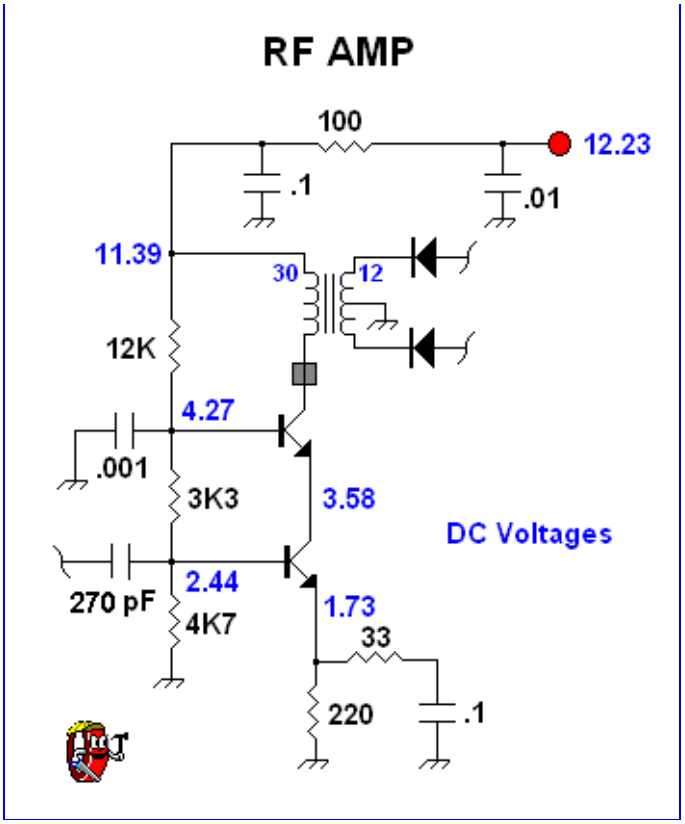
To the left is a simplified RF amp diagram taken from the schematic above indicating the measured DC voltages for reference purposes.

T1, the output transformer was wound on an FT-50-43 ferrite toroid. An FT37-43 would also be suitable. Number 28 gauge enamel coated magnet wire was used for the 30 turn primary and 26 gauge wire was wound over top to make the secondary 12 turn, center-tapped winding. The 26 gauge wire was used for the secondary winding because it provided good anchoring of the transformer by the center tapped ground connection.

You may consider substituting a 22 to 100 ohm resistor for the ferrite bead on Q1. It suppresses VHF oscillations.

Detector

To the right is a photograph of the detector from the Q1 transformer through to the JFET source follower. Schottky/hot carrier diodes or germanium diodes such as the 1N34A with a low forward voltage drop are strongly recommended. I have found there to be significant variation in



sensitivity between different types of these diodes. The 2 germanium diodes I used were matched as described on this web page. A number of detectors were built and tested for this receiver, however, the design shown had the best audio quality when compared to the others. The virtues of this detector include low noise, high bandwidth, high sensitivity and low distortion. although a little complex, this is a detector worthy of consideration in your AM receiver projects. The center-tapped Q2 transformer secondary and the 2 diodes provide full wave detection. This serves to reduce distortion somewhat and cancel even-order harmonics in the carrier signal. You may eliminate one of the diodes and convert the Q2 transformer secondary to a conventional, single link.

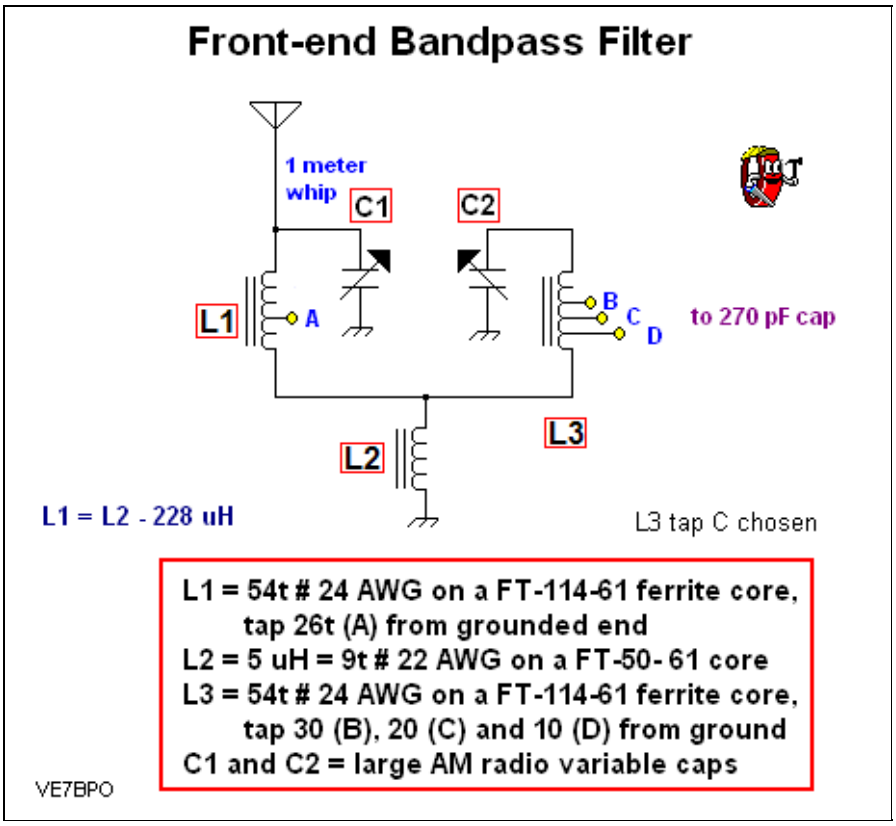
A 470K ohm resistor and R1 form a voltage divider that sets the detector bias voltage (V Bias). Some measured R1 values and corresponding bias voltages are shown in the schematic. I chose an R1 value of 100K for my final version. You may have to increase or decrease the R1 value to suit your local detector sensitivity requirements. You could also substitute a bias potentiometer for front panel adjustment of the receiver sensitivity. In this detector, changing the R1 value also changed the detector frequency response. I built a separate voltage divider with roughly the same V bias consisting of a 68K and a 15K resistor and swapped it for the 470K and 100K pair. Interestingly, the 470 K and 100K pair had better low frequency response and slightly higher sensitivity than the 68K and 15K voltage divider.
Diode detectors are best driven with a high impedance source and followed by a high impedance load. Q3, a simple JFET source follower provides a high impedance load. You might want to substitute a "popcorn" MPF102 for the high Idss J310 indicated in the schematic as a J310 is not really required here. If you substitute a MPF102, please increase the source resistor from 2K7 to 4K7 ohms.

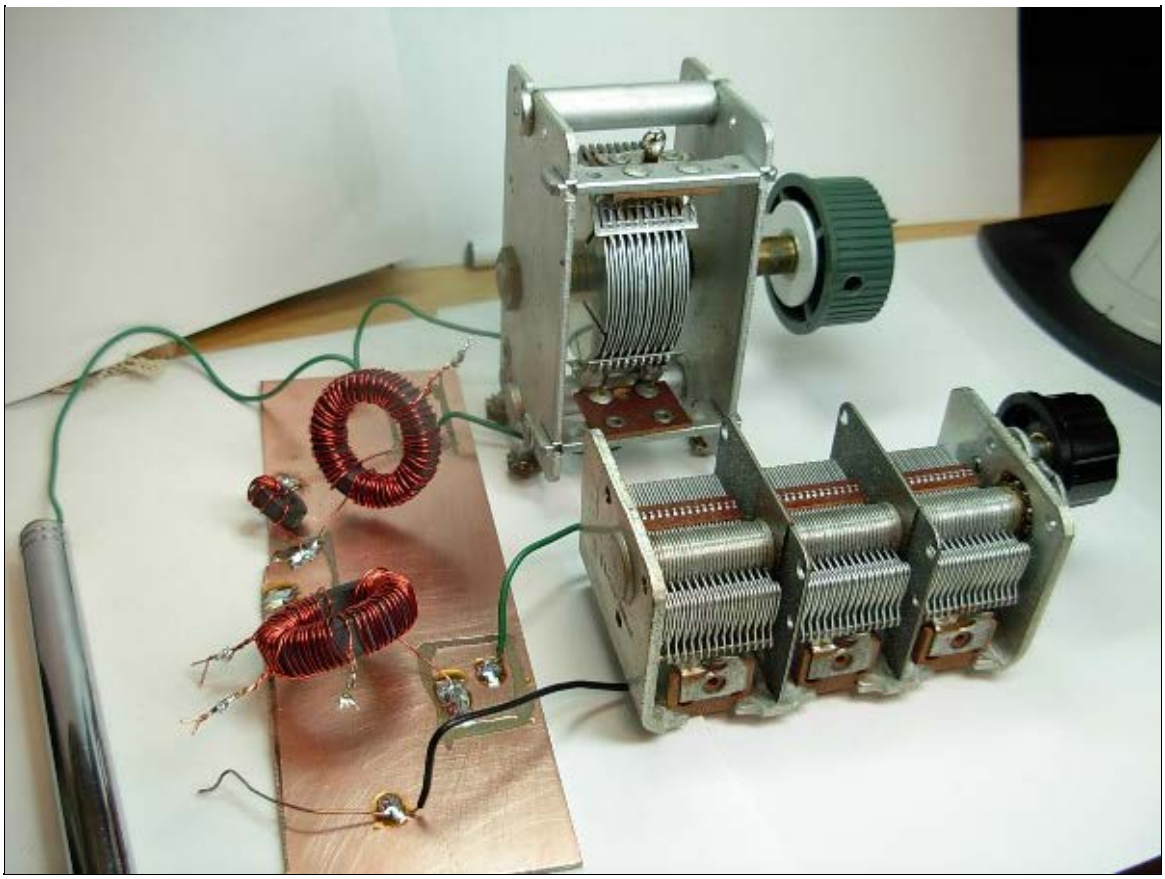
Front-end Band Pass Filter and Antenna

In late summer 2007, Wes, W7ZOI conducted experiments with zero power receivers (crystal sets and such). He wound some inductors using ferrites with an unloaded Q of over 270 at MF! Please check out Wes' web site. His work with high Q ferrite inductors illustrates the importance of quantitative measurement and also provided the following revelation; we really do not have to resort to large, air core, Litz wire coils to build high-performance inductors at MF! The early prototype front end for this project was built using FT50-61 ferrite cores, however after Wes emailed me his work on zero power

receivers, I had to get some FT-114-61 ferrites for the front end of this receiver. The next day, I emailed Mark Laurain from Amidon Associates Inc and ordered some FT-114-61 ferrite toroids. The arrival of these ferrites prompted me to finish this project and put it up on the web.

The schematic on the right is the final band pass filter used for the front end. I initially tried using just L3 for the front end, but I was unable to just tune a single station. In my city, there are 2 powerful AM radio stations at 630 and 1150 KHz. With a single inductor, I could peak one of the stations, but the other could be heard in the background. Thus, the double-tuned band pass filter presented was designed and built. Now only one station can be detected with this circuit and tuning is sharp. Most builders would use a dual-ganged variable tuning capacitor, however, I elected to use 2 separate variable capacitors. Considerable flexibility with this circuit is possible. You will have to experiment to best determine your local sensitivity versus selectivity needs and to suit the variable capacitors you have available. Large AM receiver capacitors are getting hard to find. I obtained the 2 variable capacitors shown in the photographs below from 2 old receivers found in a second hand store. One of the receivers was a Marconi tube radio that was in poor condition. I paid $5.00 for both radios and harvested the 2 beautiful variable capacitors as well as some other parts such as knobs, switches and terminal strips. Never pass up on an old, derelict radio as a potential variable capacitor source!



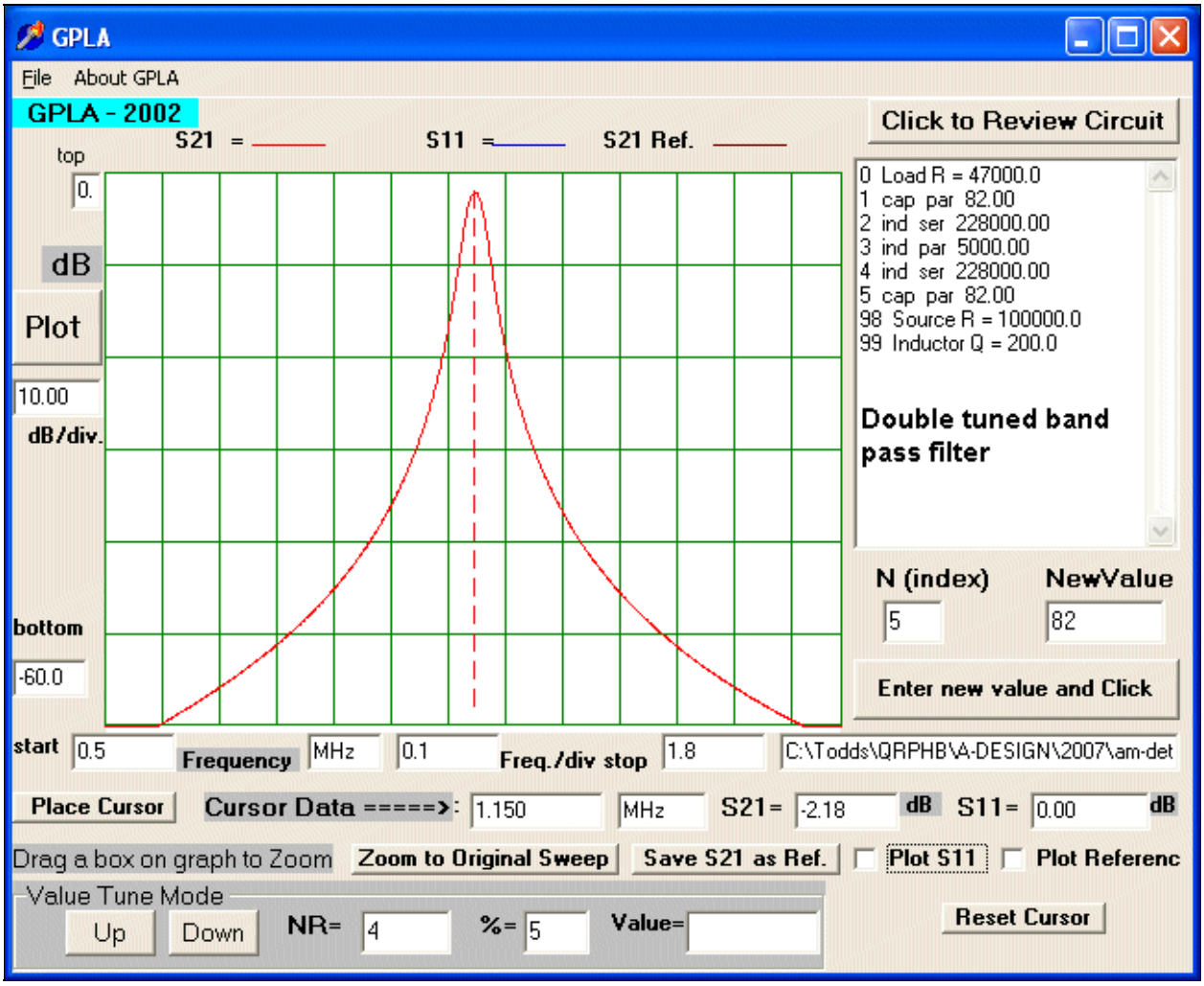


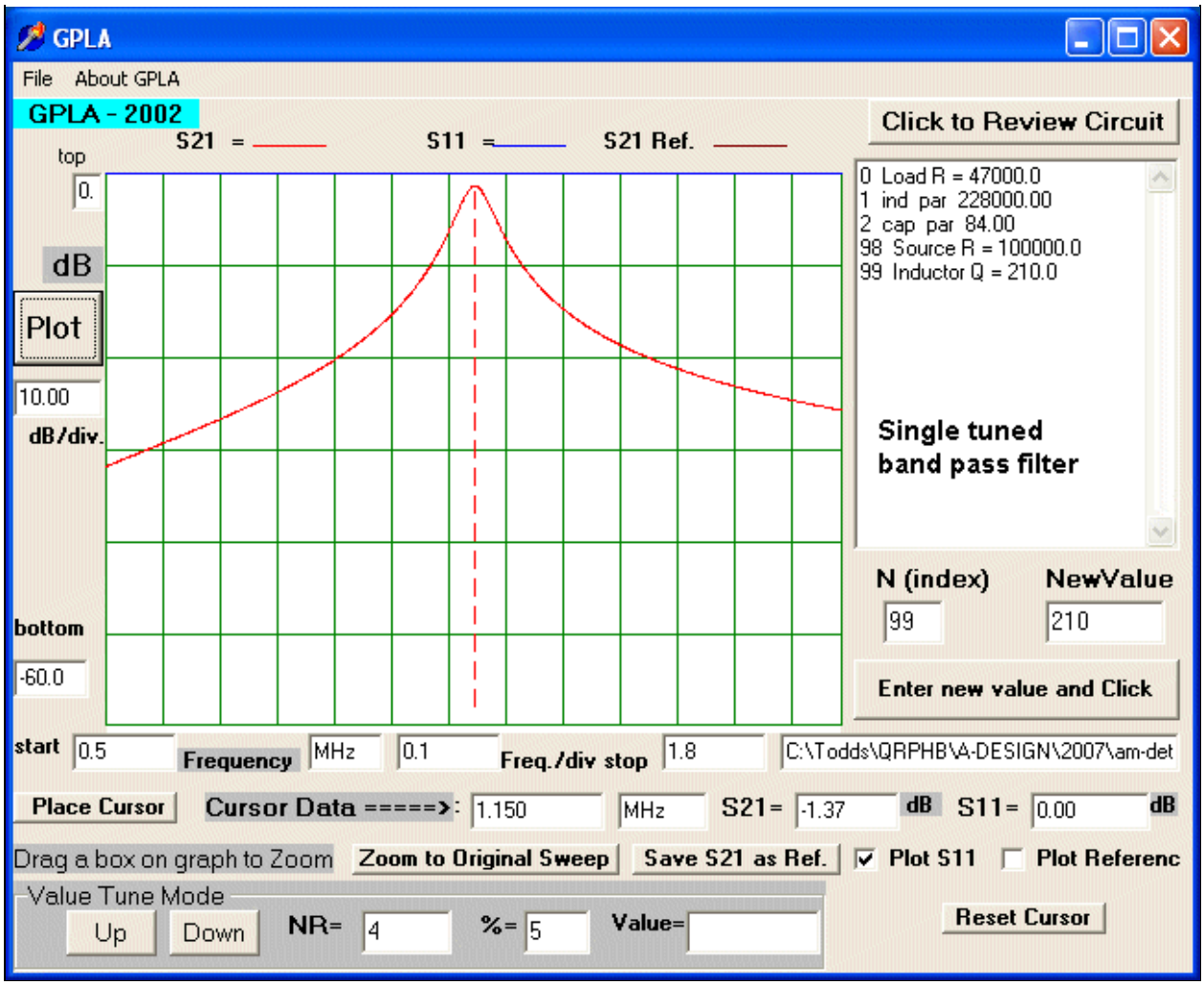
Shown above are 2 photographs of the band pass filter breadboards. The 2 variable capacitors had a variation of ~ 24 to 500 pF. There is considerable interplay between the 2 capacitors. For my QTH, it was better to peak C2 first and afterward to peak C1. Consider that L1 and L2 have a hot end and a grounded end. The antenna is connected to the the hot end of L1. Predictably, when substituting the L1 center tap as the antenna connection, the selectivity of the L1-C1 tank is increased and the sensitivity or received signal strength is reduced. This also occurs when testing the various tap points on L3 to feed the RF amplifier-detector stages. In the final circuit, I settled on Point C, 20 turns from ground. Using Point D, reduces sensitivity and increased selectivity. The opposite is true when using Point B. You the builder, have to determine which L1 and L3 connection points to use based on your own experiments and local factors. You may also change the receiver sensitivity by making changes such as increasing or decreasing the 270 pF coupling capacitor value, the emitter degeneration on Q2, or the detector bias.

Band pass Filter Analysis

It is impractical to sweep a BCB band filter using variable capacitors, so some analysis using GPLA, a program that ships with EMRFD was used to plot and better understand the double-tuned band pass filter response. A worst case inductor unloaded Q of 200 was used, but I imagine that the actual Q of L1 and L3 is much higher. For the source impedance, 100K ohm was used conjecturing that a short whip antenna at 1150 KHz would have a very high input impedance and not load down the L1 inductor. In reality, it is likely the antenna input impedance might be closer to 1 Megohm, however, I am using the worst case scenario. If the filter performs better than simulated - all is great! Higher source and load impedances and higher unloaded Q inductors would decrease the bandwidth of this filter which is desirable.
Note that I am concerned that L2 at 5 uH may may overcouple the 2 tank circuits. I did not see a double humped response on GPLA analysis, however, experimentation with L2 may be in order for the more astute homebuilder. You might consider lowering the L2 value to 3 or 4 uH and performing some testing. The load impedance for L3 was rather arbitrarily chosen. Considering that various taps on L3 may be used, the XC of the 270 pF coupling capacitor and the input impedance of the RF amplifier, I just chose 47K as the L3 load impedance. Below are 2 screen captures of GPLA plots. The top graph is the double tuned band pass filter and below it is the single tuned band pass filter consisting just of L3 and C3.
These graphs lead to 2 main conclusions:

1. The final band pass filter design appears to be reasonably sound.
2. We can understand why I could not tune in a single radio station with just L3 and C3 as the band pass filter; the filter skirts are not very steep and the second unwanted station was also amplified and detected.



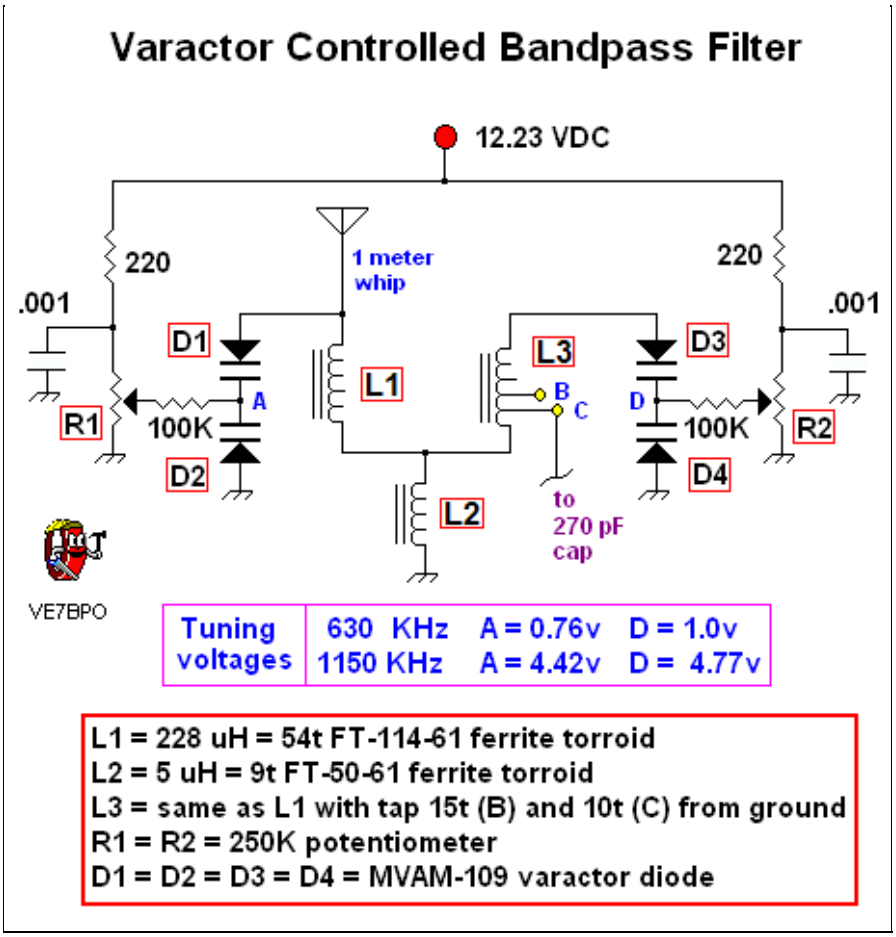


Varactor Tuned Front-end Filter

On November 11, 2007, I decided to investigate whether or not variable capacitance or varactor diodes could effectively replace the air variable capacitors in the band pass filter. In my parts cabinet were several MVAM -109 which is an obsolete but still readily available part. Another varactor, especially designed for tuning AM receivers is the 1SV149. This varactor is manufactured from Toshiba and is also appropriate.

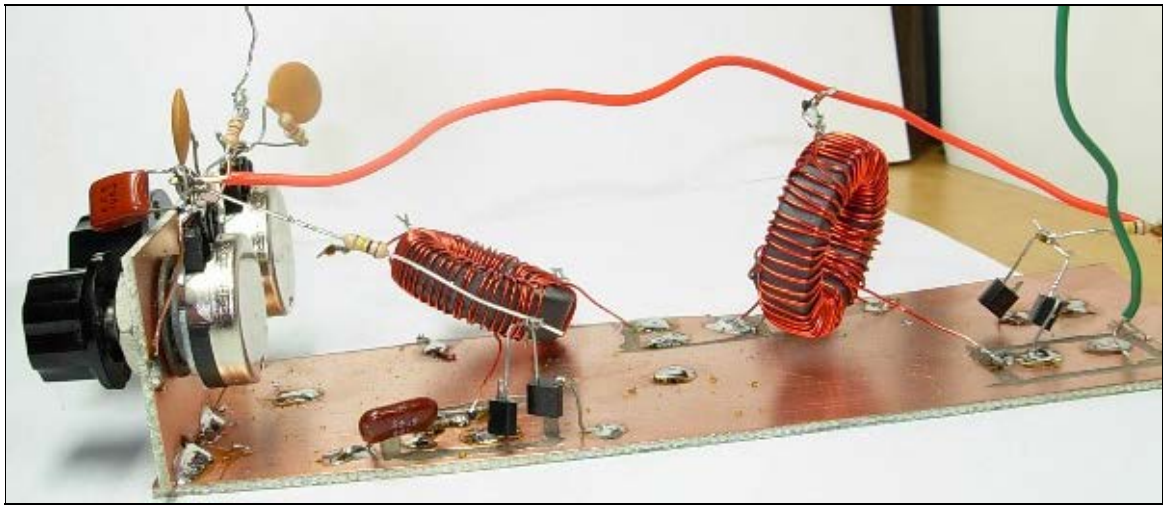
While not comparable to the Q of 300 or greater of a good quality air variable capacitor, varactors are smaller, cheaper and can be easily ganged together so that only 1 potentiometer is required to tune the front-end filter. To the right is a photo of the varactor tuned front-end filter breadboard.

To the left is the schematic of the varactor tuned front-end band pass filter. The air variable capacitors were unsoldered from the original filter breadboard and a small board drilled and fitted with two 250K potentiometers was soldered to it.



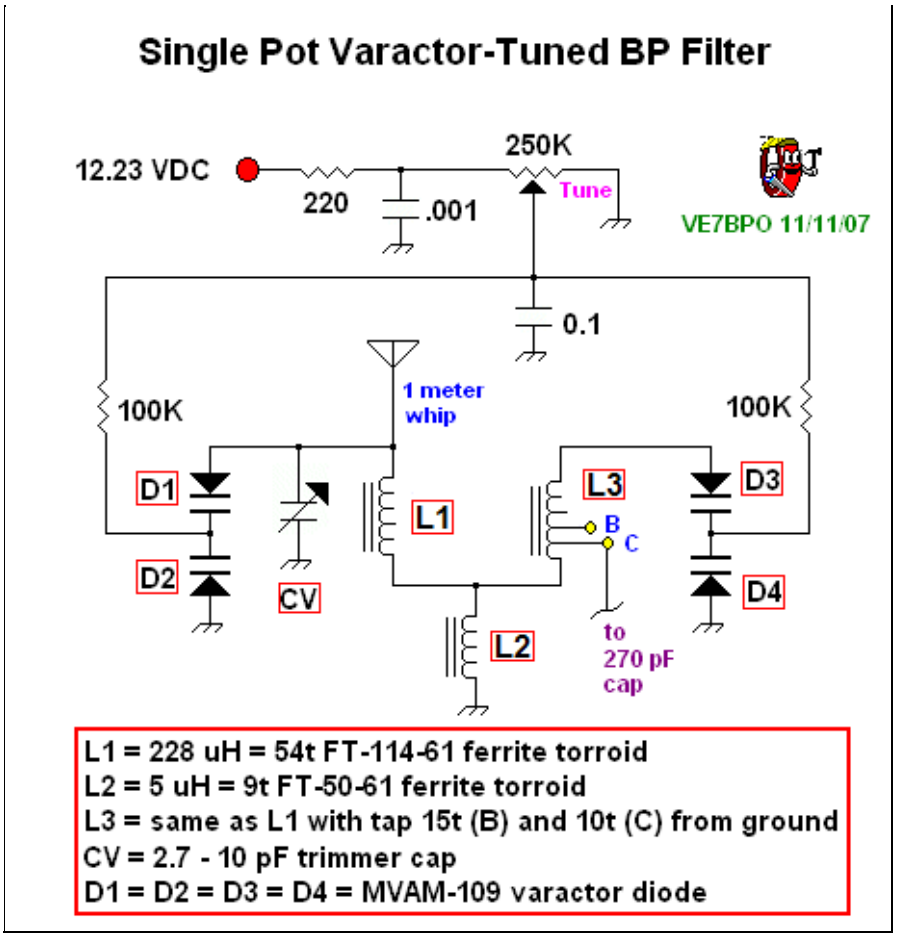
L3 was also modified to have taps at 10 and 15 turns from ground. I conjectured that since the varactor diodes have less Q than their air variable cousins, it would be wise to tap down on L3 to try and increase the selectivity of the L3 tank circuit. In the end, I used the tap at 10 turns from ground for my receiver as signal strength was still acceptably strong. You may choose to use the tap at 15 or some other point to suit your local selectivity/sensitivity requirements.

I was able to tune in single stations as I previously did with the air variable capacitors. Tuning is "touchy". Ten-turn pots would be a better choice, however, are not very frugal for such a project. You get used to tuning with conventional potentiometers after a few minutes or so. I measured the reverse voltages required to tune the 2 main local AM radio stations and they are tabled in the schematic. The L1 tank requires slightly more capacitance to resonate than the L3 tank. Thus it takes a little less applied reverse voltage to the varactor pair resonating the L1 tank compared to the varactor pair resonating the L3 tank .A side view photograph of the varactor breadboard is shown directly below. The component leads have been kept long so that I can recycle parts from experimental project to project as possible. This helps contain costs. Shorter lead lengths and proper lay out should be pursued in any final projects you build.



Single Varactor Tuned Front-end Filter

Tuning with a single potentiometer ganged to both varactors is easy to do after learning from the experiment above. All that is required is a method to compensate for the differences in capacitance between the the 2 LC tanks. I placed a high-Q (Q=300) variable trimmer capacitor in parallel with L1. By listening to the receive signal strength and tuning in one radio station using the potentiometer, I was able to peak CV for the strongest signal. I did this for both 650 and 1150 KHz and actually unsoldered CV and measured its value with a capacitance meter. The CV value was ~ 6 pF for both frequencies. I decided to replace CV with a fixed 5 pF silver mica capacitor and left it there in my final filter version. Your results will probably be different. I suggest just leaving CV



and using this trimmer cap to peak the signal once you have tuned a desired radio station with the main tuning potentiometer. An alternative to using CV is to vary L1. You could try compressing the number of L1 windings to allow tracking of the 2 LC tank circuits.

For the varactors, I used back-to-back VVC diodes as opposed to just a single varactor to resonate each tank. This was done in an effort preserve the highest varactor Q possible. The RF voltage of the AM RF signal may be high enough to forward bias a single varactor during a portion of the AC signal and degrade Q. This does not happen when back-to-back diodes are used. Almost all high-grade FM tuner schematics I have seen use back-to-back varactor diodes in their various ganged, tunable band pass filters. The major drawback of back-to-back diodes is your tuning range is reduced because you now have 2 capacitors in series. Experimentation may be required to achieve the BCB band-spread that you desire. You can add another pair of varactors in parallel or add some parallel fixed capacitance or even change the L1 and L3 inductance values for example.

This receiver tunes nicely and sounds fabulous. Last evening I was able to tune in 5 different AM stations, however, other than the local 2 radio stations, the others were quite faint. This is not bad considering this receiver has only 1 RF amp and a 1 meter long antenna. This band pass filter could be adapted as a pre-selector for AM radio reception. To match 50 ohms, lower L1 and L3 tap points could be chosen or a few links of wire may be wound around the inductors.

In the photograph below, you can see the 5 pF capacitor soldered in parallel with the MVAM-109 pair associated with L1. The antenna also connects to the ungrounded end of the 5 pF capacitor. Below in the last photograph; since only one potentiometer is used for tuning, a large knob was screwed on to the pot control shaft to make tuning a little easier. The solder-laden 220 ohm resistor is the connection point for the regulated 12.2 VDC. The 220 ohm resistor on the left has been cut from the 12.2 VDC connection point so 0 voltage goes to the left potentiometer.

Final Thoughts

I emailed Wes, W7ZOI and asked him why it is better to inductively couple a tuned circuit which use air variable tuning capacitors. Wes wrote his answer in the form of a complete web article entitled Coupling Methods in the Double Tuned Circuit. Big thanks Wes! From his summary, when the inductors used to resonate each tuned circuit are constant, and inductive coupling is used, the coupling of the resonators will remain constant as the variable capacitors are tuned across the band. Please download and study his web article for it not only discusses coupling in the double tuned circuit, but provides some insight into using his LadBuild and GPLA software from EMRFD.

Junk Box NDB Low Pass Filter

Introduction

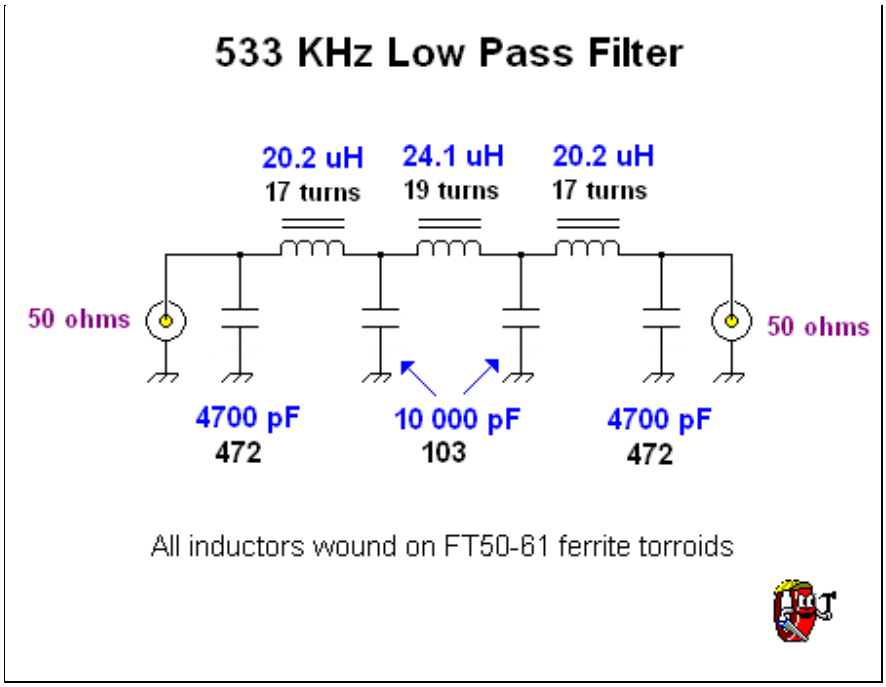
With winter approaching, many HAM and SWL hobbyists find intrigue in tuning in NDB or **N**on-**D**irectional **B**eacons. Although the tunable NDB band depends on your location, in Canada it may be found in a band ranging from about 190 to 535 KHz. Canadian beacons either have just a carrier (no offset)or are tuned using the USB with about a 400 Hertz offset, however, different offset frequencies and certainly LSB are used when receiving DX from other countries.

Less than 10 Km away from my QTH is a 10 KW AM radio station at 1150 KHz. On my test receiver, the S-meter reads off the scale (> 60 dB over S-9) when tuned to this frequency. This local radio station causes spurious, second-order intermodulation products (direct mixing) that all but wipes out some weaker NDB stations that I am trying to tune in. Certainly, having a 500 Hertz crystal IF filter is useful, but attenuating this local QRM is also desirable and is the topic of this web page. Many general coverage receivers offer limited or in some cases no filtering of the NDB band, however an outboard filter is an easy project to build in one afternoon.

Update Oct 11, 2010: Here is a link to a version of this project built by Robert, K5TD

Project Schematic

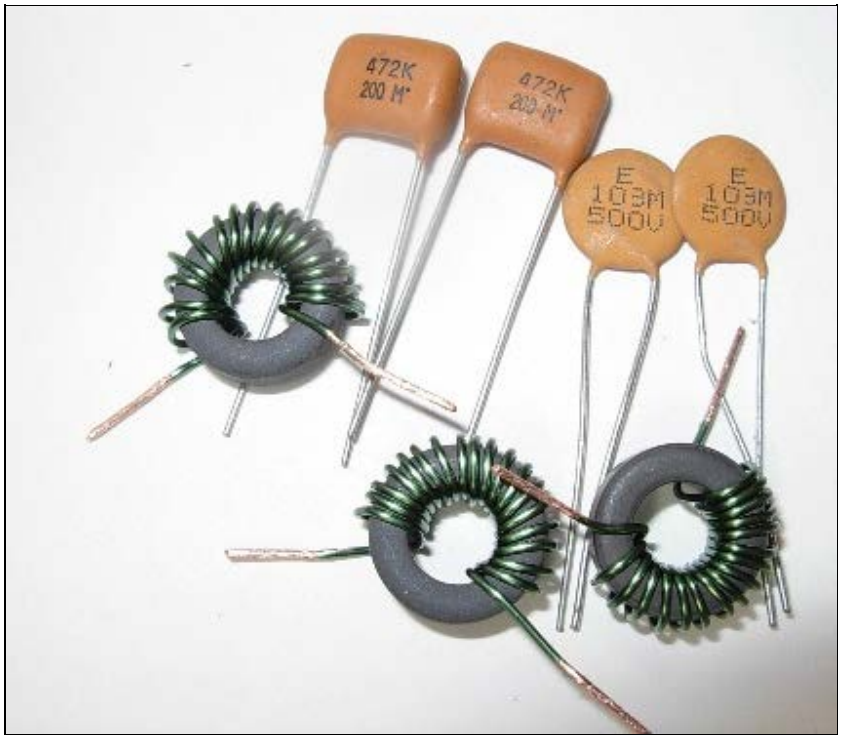
To the left is the project schematic. It seems odd to build a low pass filter to reduce BCB interference (as usually a high pass filter is required for this purpose at HF) however for NDB, an aggressive low pass filter is required. For simplicity sake, a 7 element Chebychev low pass filter was chosen. Since it is easy to wind reasonably high-Q inductors for 10 uH and greater inductance using number 61 material on a ferrite torroid, the FT50-61 core was chosen for all of the inductors. Number 22 gauge wire was used for the coils to keep the unloaded Q as high as possible. The FT37-61 ferrite is also suitable, but will have less Q and require smaller gauge wire. Use 19 turns instead of 17 for the 20.2 uH and 21 turns instead of 19 for the 24.1 uH coil. Do not use number 43 material ferrite cores.



Components

I do not stock RF capacitors greater than 2200 pF, so junk box ceramic capacitors were used to build this filter. In fact, this design specifically uses more common, standard value capacitors to reduce cost and to not have to order in parts. Certainly, the astute builder could use higher quality capacitors or even large powdered iron torroids instead of the ferrite cores for inductors if higher performance is desired. Try to use high Q capacitors if you can find or are purchasing them. Poly or silver mica caps would be great choices. You can substitute a 5000 pF capacitor for the 4700 pF called for in the schematic.

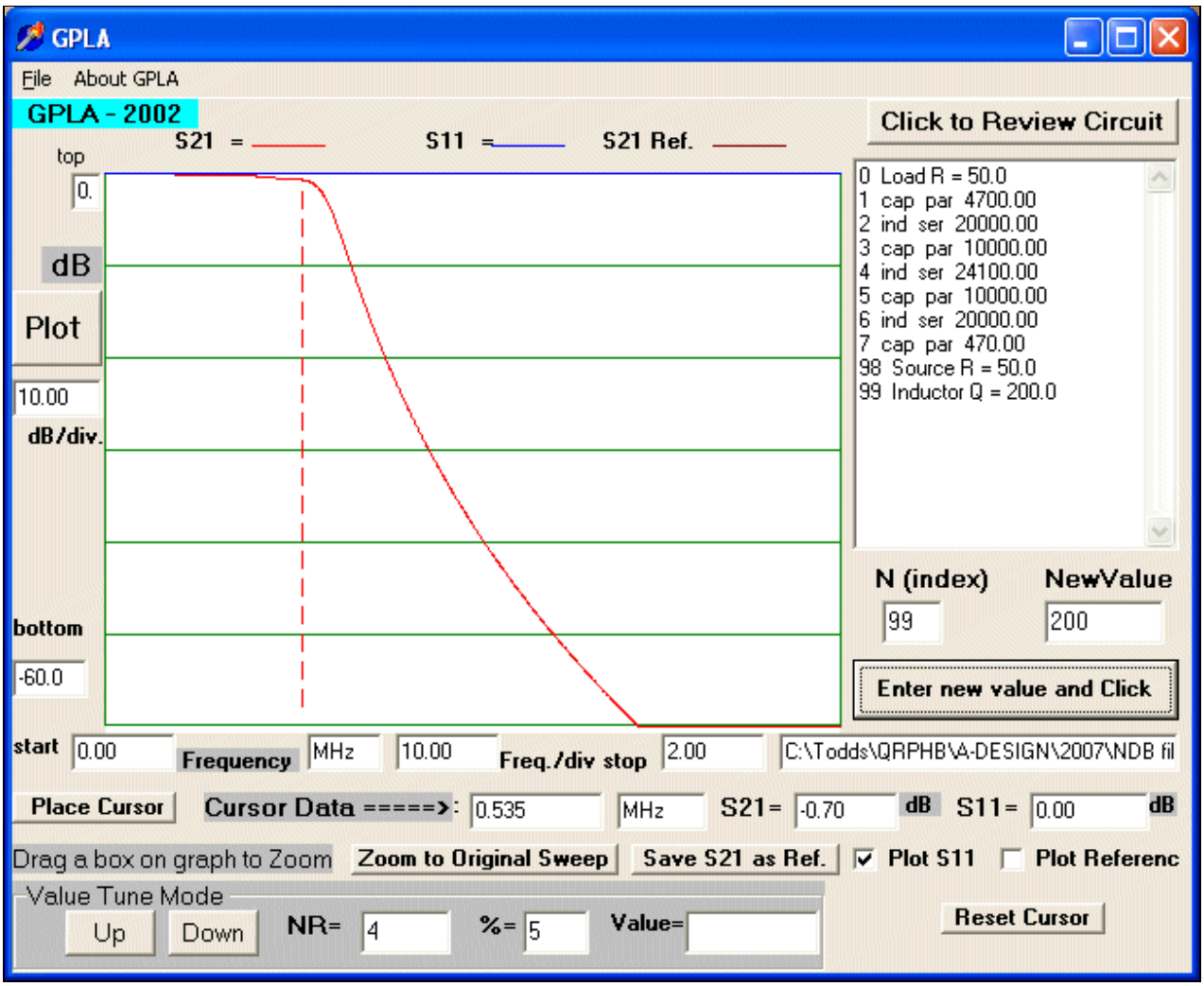
To the right is a photograph of the components I used in the project breadboard.



Breadboard

To the left is the completed project. Ugly construction as always, was used. The inductors were spaced at least 2.5 cm (1 inch) apart at right angles to try to minimize unwanted coupling.

GPLA Simulation



Above is the plot of the filter during simulation with GPLA. The simulation calculated an attenuation of ~46dB at 1150 KHz. At 1000 KHz the signal was 40 dB down, at 800 KHz it was ~24 dB down and at 630 KHz, the attenuation was only ~5.7 dB! Clearly this filter is not suitable if the offending BCB interference is from a station significantly less than 0.8 MHz. For my situation, this filter is acceptable. A 5 element Chebychev filter was also designed and plotted but was discarded as there was only 32 dB attenuation at 1150 KHz.
Since I wanted to tune as high as 535 KHz, the 533 KHz cutoff frequency was chosen. Additional work to help those with strong BCB interference at the lower BCB will be attempted in the future and presented on another web page.

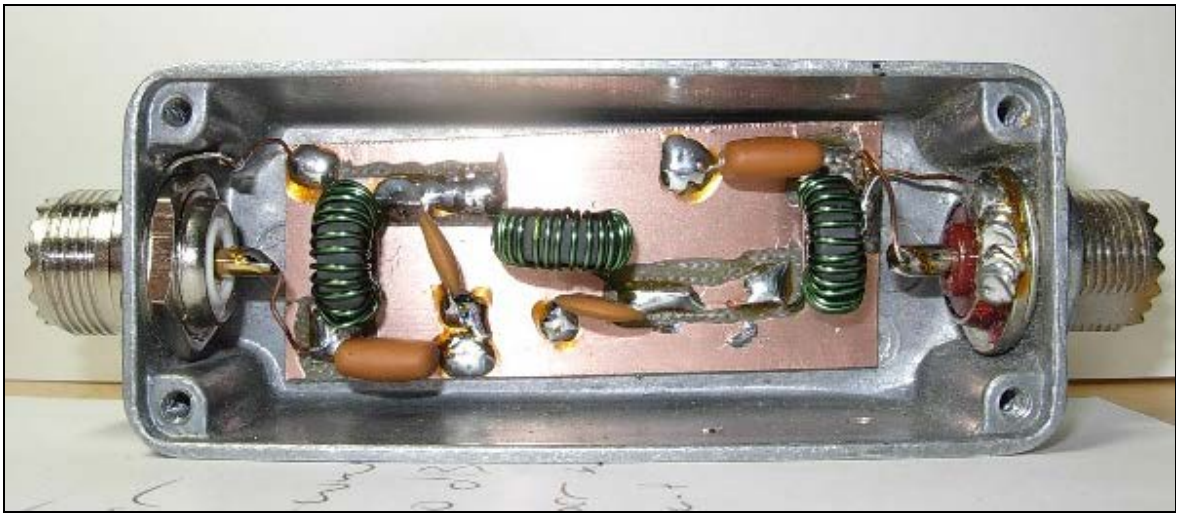
Receiver Testing and Comments

Click on the picture to the left to hear the beacon YWB at 389 KHz with a 500 crystal hertz IF filter engaged on a borrowed Icom R-75 receiver. W7ZOI did some measurements on his R75 receiver S-meter using a signal generator and step attenuator. From S9 on up to 60 over, the steps were very accurate. However, below S9, correlation was poor. The built in attenuator is -20 dB when engaged. On my test receiver, I did some A/B testing with the filter in or out. For 1150 KHz (without the low pass filter) I had to engage the receiver's attenuator as without it, the S-meter reads off scale. With the attenuator engaged, the S-meter reads 50 dB over S-9 when tuned to 1150 KHz.

With the filter connected between the receiver and the feed line, (and the attenuator engaged) the S-meter read S-9.This is a drop of about 50 dB at 1150 KHz which means that this filter pretty much works as designed. I love the Icom R75 receiver; it is good value with it's many features in a compact package. Further testing was undertaken on other frequencies.

When listening to **WWV**; At 2.5 MHz, without the filter, the S-meter read S-9. With the filter inline, I could not hear WWV or see any S-meter reading. At 5 MHz without the filter, the S-meter read was at 20 dB over S-9. With the filter inline, I could still hear WWV very faintly, but the S-meter did not register. There was little noticeable attenuation at less than 700 KHz when using this filter.

Additional Photos and information



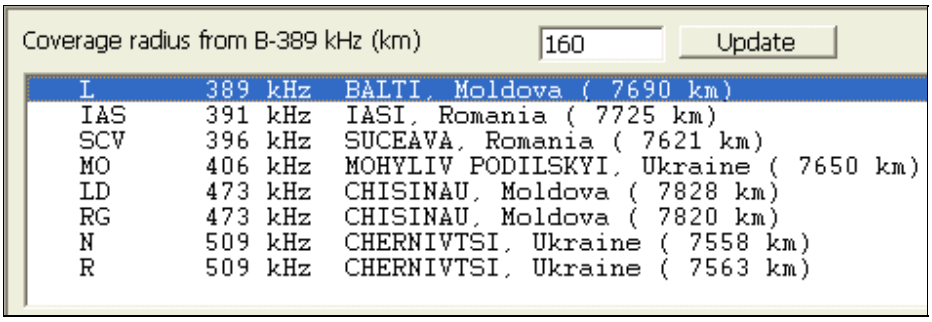
Shown above is the completed project in a Hammond die-cast case with SO-239 connectors at each end.

Shown above a photo of the Skookum beacon SX. It is on 389 KHz. This NDB is located in Skookumchuck BC, Canada

Links

My friend and fellow NDB enthusiast, Ken, VE7KPB has a posting on his web site showing some of the beacons he has logged from his QTH. Consider trying some of these frequencies from your own QTH to get used to finding beacons. Note you must temporarily allow pop ups to see his excellent log.

We recommend this non-directional beacon search and log utility program called WWSU from VE3GOP It must be registered and is a wonderful low cost tool.
I was near beacon L in Balti, Moldova (Балти, Молдова) in 2006. Below is a snippet from the VE3GOP program showing beacon L and also some nearby beacons. (Я изучаю русский язык). Приветствую Вас дорогие друзья!



Martin Francis has an excellent NDB web site including the free program called NDB WEBLOG for a number of platforms

Some beacons may be located using this NavAid web site

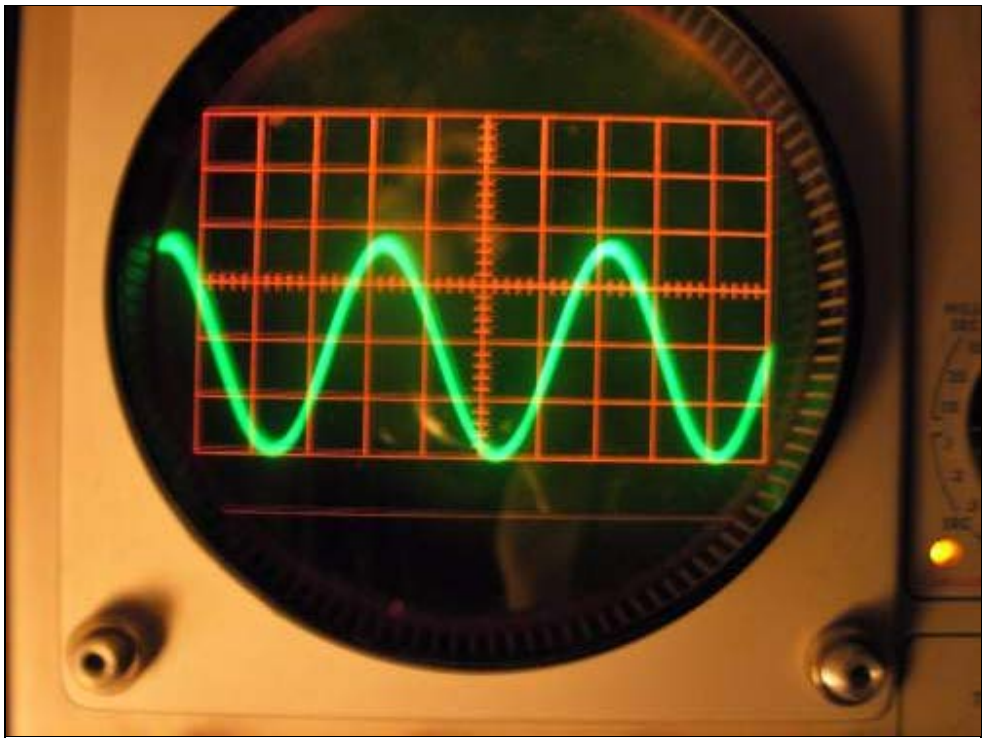
Clint, KA7OEI has an informative web site regarding NDB listening including using digital computer processing to dig out weak signals. This is also a great overview site for newcomers to NDB.

Conclusion

To the right is the outcome when I connect my frequency counter directly to my antenna coax cable feed line. 1150 KHz is my nemesis frequency! Happily it can be tamed with a little filtering to allow NDB listening and logging.

To the right is the outcome with my SWL antenna coaxial feed line connected directly to the scope. The scope was on the 0.5 volts per cm scale, so the peak to peak voltage is 0.2 volts. In just about any high-gain audio amplifier I build, if I touch my finger to the input, I can hear AM 1150 loud and clear - no wonder!

Good luck with your own NDB adventures!



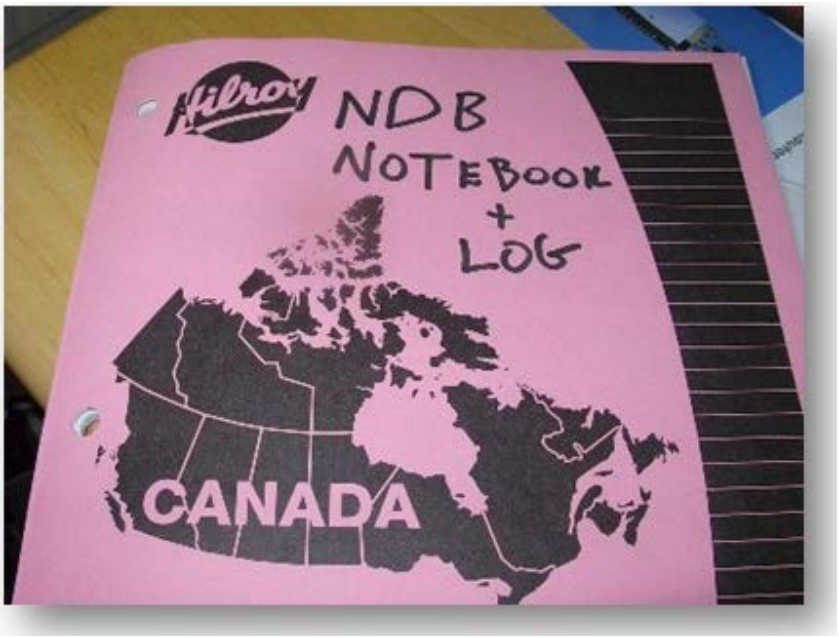
More NDB Information and Circuits

Introduction

Latest Update: December 3, 2012

This web page holds a collection of NDB-related ideas, experiences and hopefully will include some feedback from fellow NDB enthusiasts.

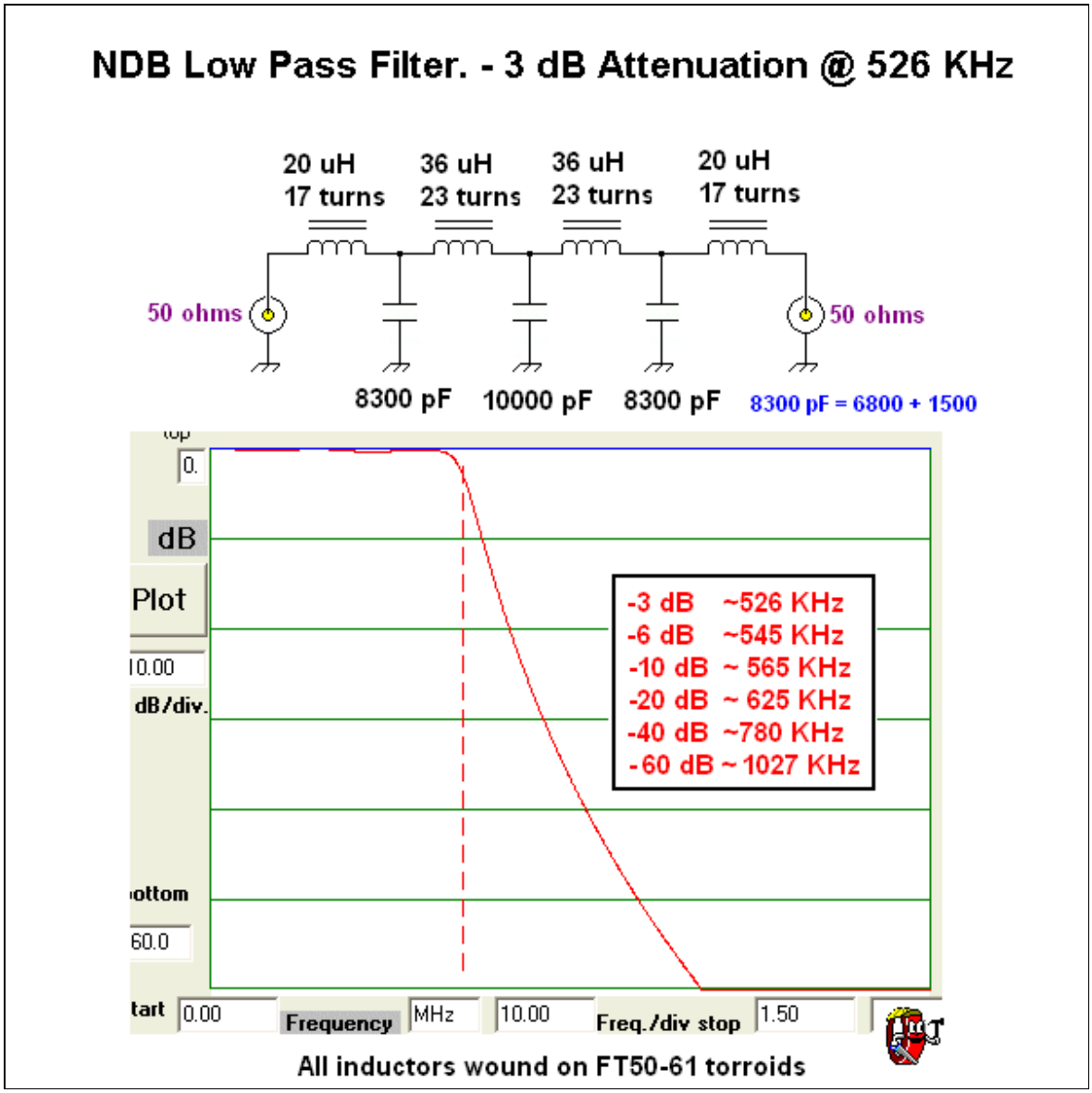
I devoted a new notebook to this topic and hopefully with inspiration from band listening and communicating/learning from others, I will fill it over time. New content will be added to the bottom of the existing material as QRP-Postadata



Improved NDB Chebyshev Low Pass Filter

A popcorn or "junk box" low pass filter was designed and presented on this web page. After discussion with VE7TW and testing a Realistic DX-300 and other receivers, it became apparent that even more attenuation of a strong local BCB station at 1150 KHz was desirable. In addition, there are other moderately strong AM radio stations from 630 to 800 KHz (especially at night time) which maybe causing mixer intermodulation distortion products. A fault of the junk box low pass filter is poor attenuation below 800 KHz and a better design was a prudent goal. Building on the learning obtained from the junk box filter experiments, an improved 7 element Chebyshev low pass filter was designed and is presented directly below. The 3 dB cut off of this filter is calculated to be 526 KHz. This is the filter that I now use for my home radio station. At my

nemesis frequency of 1150 KHz (where a powerful local radio station broadcasts), the attenuation is calculated to be 68 dB. It takes careful layout and a conductive chassis to realize this level of attenuation, but the effort is worth it. In very strong AM BCB locations, you might consider placing 2 such filters in series between your antenna and receiver if required.



The schematic and simulation of the improved NDB low pass filter is shown above.

Non-directional Beacon Identification

It is interesting to visit nearby beacons. In the photograph to the right is XC which broadcasts at 242 KHz.

I have learned that it is very important to confirm the NDB stations your hear via a database or list. What you hear on the air should match the database/list for **both** call sign and frequency, else suspect that you may have copied it incorrectly. **RNA**, the definitive signal list for North and Central America plus Hawaii may be found here.

Three Questions

Steve Ratzlaff, AA7U is an experienced NDB DXer and has been listening to beacons since the mid-1980's.
I asked him the following 3 questions:

1. LF beacons do little more than send their station identification in Morse code, are mostly low power and generally might be perceived by some people as boring and low tech. Yet, on the World Wide Web, one finds numerous web sites, software, projects and commercial equipment all passionately dedicated to NDB listening. What's all this fuss about listening to beacons?

Steve: It's a hobby that requires quite a bit of skill and technical accomplishment to get the most from the equipment. Most folks have AC noise to deal with, which can be particularly bad at LF. Finding an antenna that works at LF and that can be used at your own location can be a major task; finding a radio that has decent LF sensitivity, or an LF converter to use with an existing radio--all these must be detitle with just to begin hearing anything at LF. I find it to be quite a challenge. If it were easy to receive LF beacons then I probably would have lost interest years ago! It's true that in recent years several software programs have become available that allow finding beacons somewhat easier--one simply looks for them on the computer screen and decodes the dots and dashes of the beacon being received. This is quite popular among beginners and veterans alike. But the traditional method of aurally listening for the morse code idents of beacons is probably used more often, though many are combining both aural and software techniques now.

2. Let's say I live in a small city lot or even an apartment. I have modest equipment and/or not a huge amount of cash to spend on gear for NDB listening. From the antenna through to the headphones, what are some basic recommendations you might give to a newcomer wanting to get started in NDB listening?

Steve: The radio must have decent sensitivity at LF, or else an LF converter must be used. Due to high local AC noise, any type of LF antenna used indoors will be a poor substitute to one that can be placed outdoors. A few portable radios cover the LF NDB frequency range that will work for hearing local beacons, though the radio may need to be used outside to get away from AC noise. The discontinued Sony 2010 was the standard for portable radios for reasonable LF performance. Newer radios like the Degen DE1103 have been found to work reasonably well at LF and can be bought for well under $100 by mail order from eBay sellers; or the more expensive Kaito 1103 version, which has a warranty, can be obtained from several distributors like Universal Shortwave. The much more expensive semi-portable Eton E1 works well at LF, but is more in

the price range of a tabletop radio. The Icom R75 is currently the best bargain in a tabletop radio that has very good LF sensitivity as well as 1 Hz tuning, which is an asset if a narrow external audio filter is used. I'm not too optimistic about what someone living in an apartment or high rise building might do to successfully receive LF beacons indoors. Often the AC noise level is too high to be able to use an indoor antenna. But some have been able to use loop antennas indoors for the stronger signals. An example of a top of the line commercial loop would be the Wellbrook ALA1530 or LFL1010. Unlike at shortwave frequencies, where simply tossing a wire out the window to a nearby tree or other support, or even running the wire around the room inside, will usually work fairly well, at LF a wire less than several hundred feet generally doesn't perform very well. It can be argued that an active whip antenna makes a very good LF antenna, and doesn't take up much room, but it must be used outdoors. And if there are strong AMBCB signals, then the active antenna, either loop or whip, must have very good overload resistance otherwise it can generate distortion of its own from the strong BCB signals.

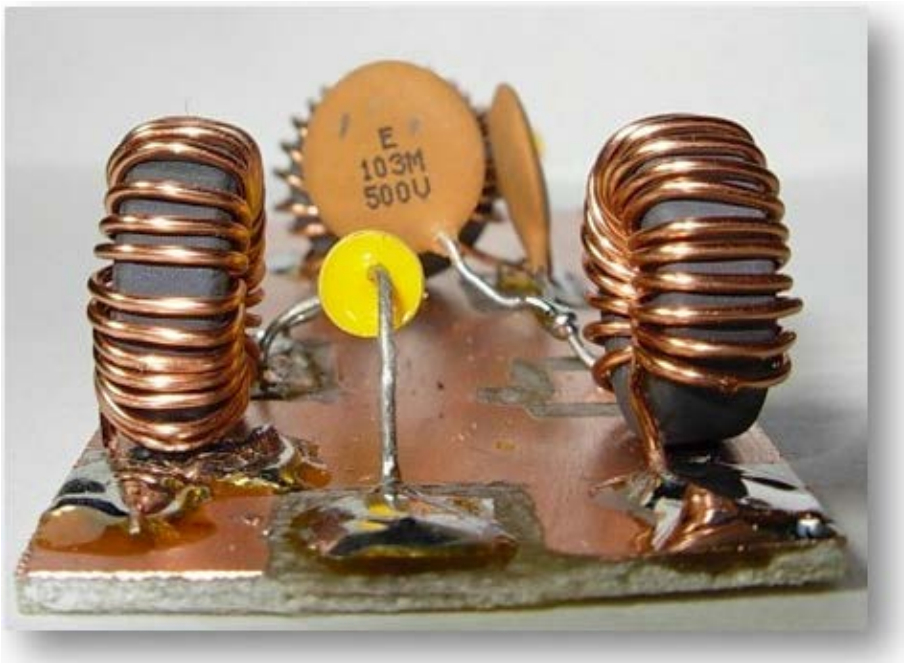
3. What kind of distances are considered DX for NDB?

Steve: NDB DX is pretty much a relative thing. One just starting out might be thrilled to hear a beacon from the next town, or from the other side of his own state or province. As one improves his listening setup and gains experience, then usually DX goals also expand to try to hear beacons farther and farther away. NDB DXing generally is not a competitive hobby, unlike amateur radio with its various competitive "contests". Each person's listening setup, local noise level, etc. is usually very different from someone else's, even someone in the same town or general area. One person might live in the suburbs and have a lower noise level than his friend who lives right in town and has a much higher noise level. One might have room to put an antenna in a quiet spot; the other might be limited to much less. People who live near an ocean generally have a much better chance at hearing something exotic offshore than folks living far inland. Folks living in the central or eastern part of North America have many more beacons available to be heard than folks in western North America. But there are always a few beacons that are much stronger than most, and can be heard from long distances of 1000 miles or more, pretty much anywhere in North America at night. One example would be 206 GLS in Galveston, Texas, which runs around 2000 watts, has a large antenna, and is generally readily heard anywhere in North America at night--that beacon might be 1500 miles or more away, and might be considered real DX. However another 25 watt beacon from the same general area in Texas might be hard to hear only several hundred miles from that beacon. So "DX" is pretty much a relative term. Ndblist, an international email list devoted to beacons, is open to anyone with an interest in beacons--members post their loggings there. What might be a local beacon to someone might be DX to someone in a different part of the country. All levels of experience are welcome.

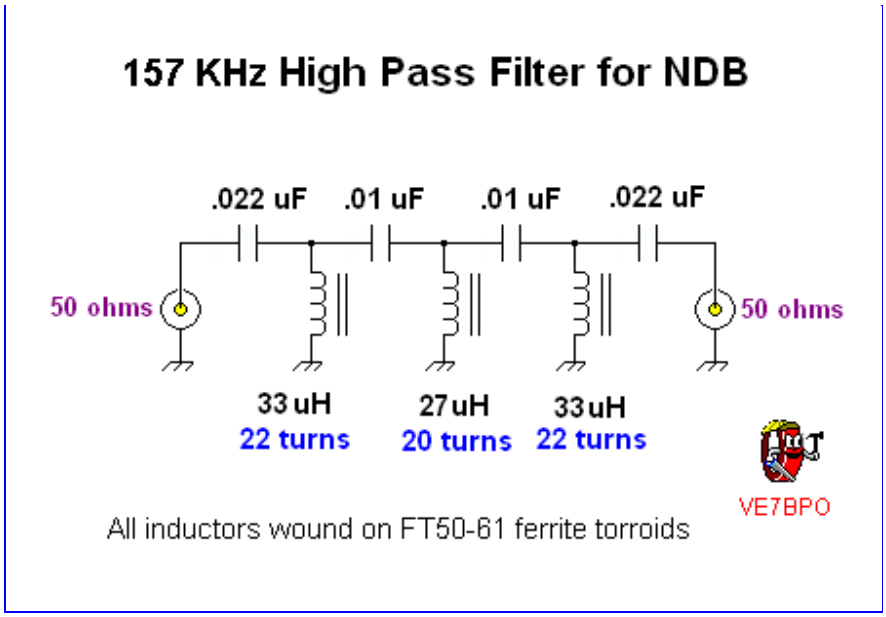
Thanks Steve.

NDB High Pass Filter

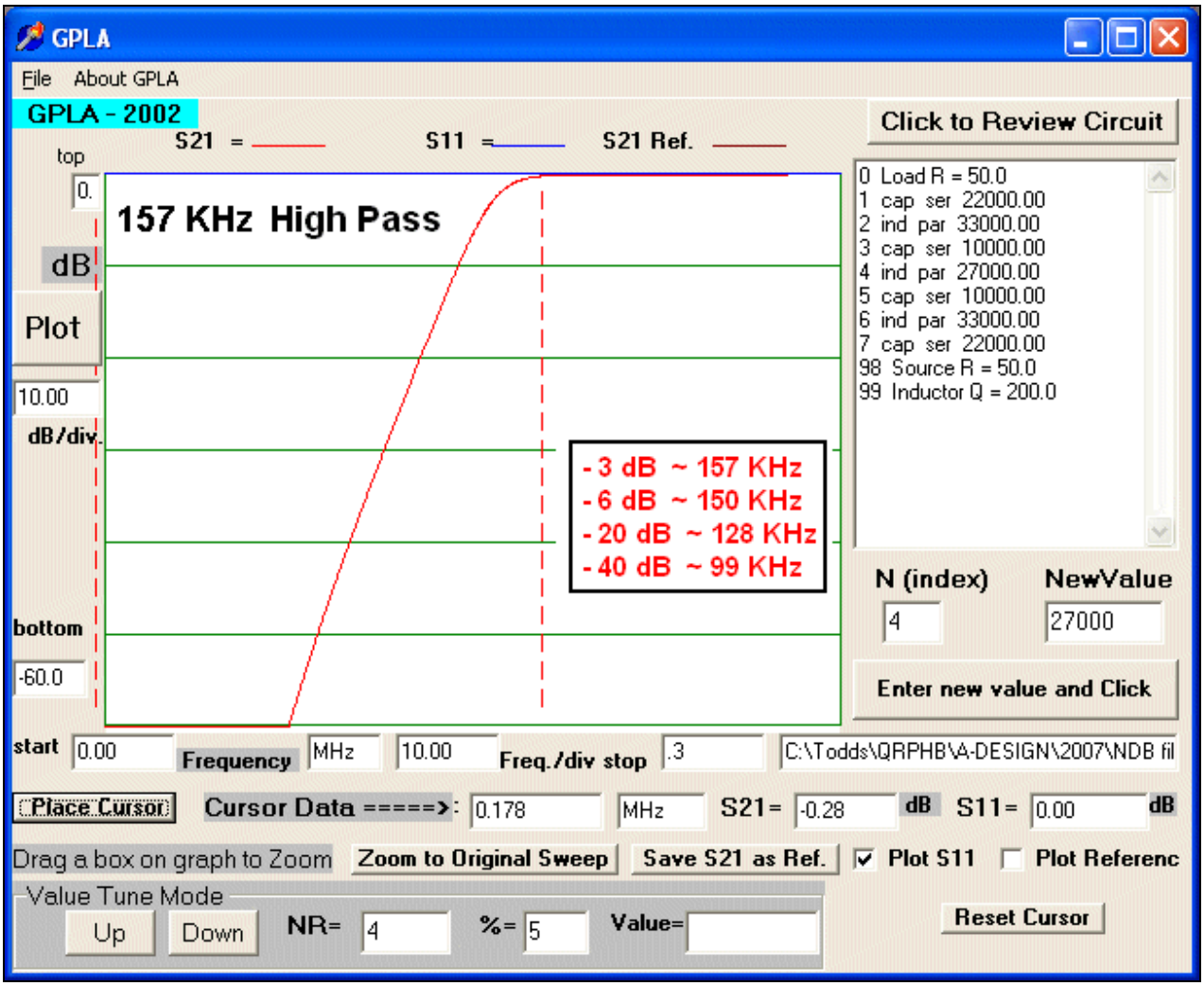
A high pass filter using standard value capacitors was designed using GPLA. although, such a filter would not help AC line noise and RFI generated in the house, I suspected my antenna was picking up local noise from below the NDB band. This filter was mounted inside a die-cast Hammond box with a SO-239 at each end. I used 22 gauge enamel covered wire for the inductors. A photo of the filter is shown to the right.
For the 0.01 uF caps, I used junk box ceramic capacitors with a 20% tolerance, however, I did measure a bunch and found 2 within 5% tolerance for my filter bread board.



To the left is the filter schematic. This is an N = 7 Chebyshev high pass filter with a 3 dB cut off of 157 KHz. This cutoff frequency allowed the use of common, standard value capacitors and also even turns numbers to reach the desired inductance for the inductors when wound with FT50-61 ferrite cores.



Use 5% tolerance, high Q caps such as polystyrene or NP0 ceramic and **not** junk box bypass-grade ceramic capacitors as possible. I used trashy ceramic caps for the 0.01 uF parts due to lack of better parts at the time of building and testing.

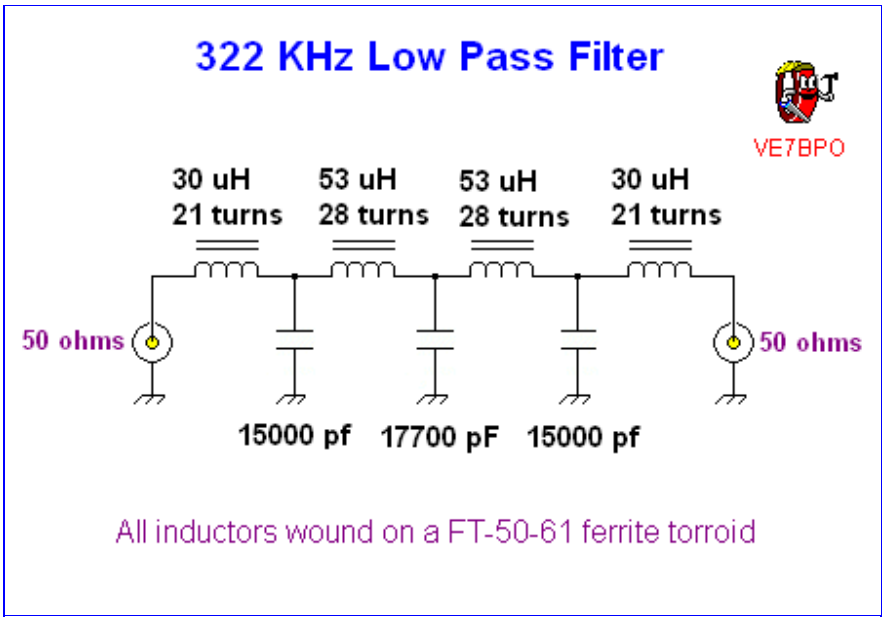


Above is the filter GPLA simulation. In particular, I have harsh noise from about 110 KHz on down. At 78 KHz, where this filter has a calculated attenuation of ~ 56 dB, I made an audio file of the band noise. This is in AM mode with the filter out for a few seconds and then in line. With the high pass filter in line, there is pronounced attenuation of the noise and my local 10 KW BCB station at 1150 KHz suddenly appears. Prior to this it was hidden by the harsh noise. At frequencies less than ~200 KHz (without my low pass filter) I can hear this BCB station intermittently as I tune around. I suspect that the R75 filtering down at 200 KHz and down is insufficient to stop this monster station.

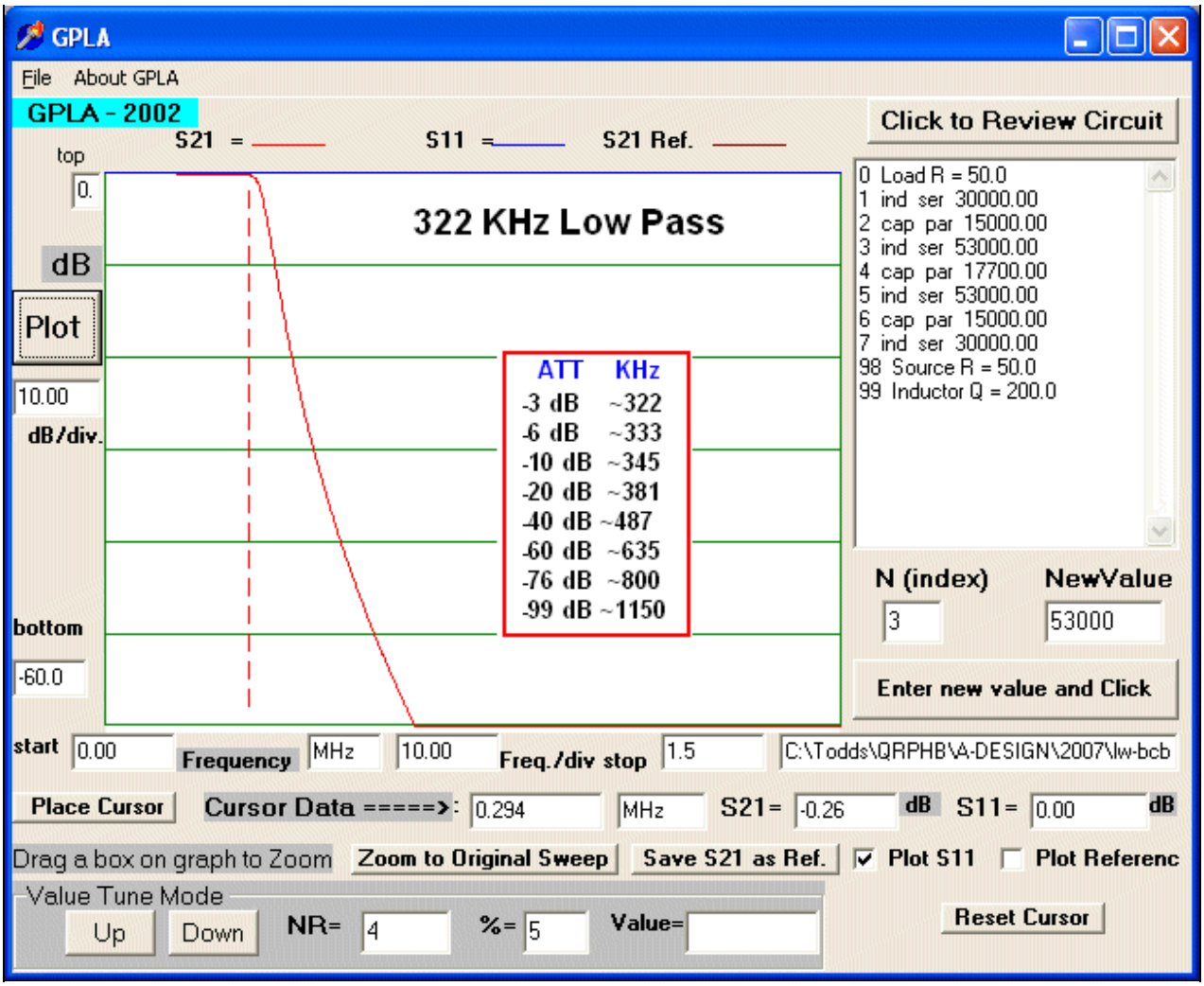
At my QTH, using a high pass filter reduces some of the noise on the NDB band. At my location, a high pass plus a low pass filter in cascade between my antenna and my receiver results in less QRN and easier weak signal copying.

Long Wave Broadcast Radio Filter

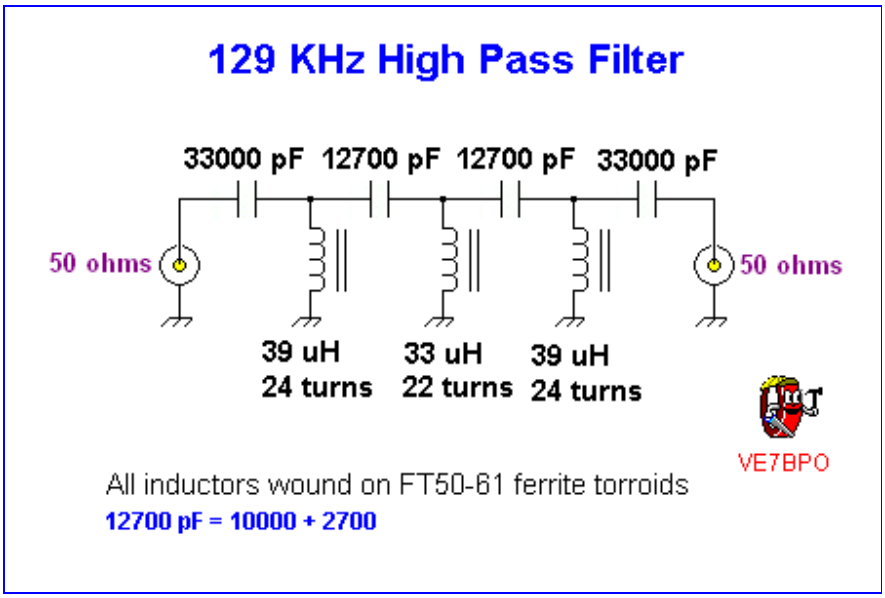
I learned about LW Broadcast radio from Steve Ratzlaff. In particular, радио россий "Rah-deo RaSEE" (make sure you roll the R!) can occasionally be heard on the west coast and broadcasts at night-time using 500-1000 KW power. The frequencies he recommended to try were 153, 180, 189, and perhaps 171, 234 and 279 KHz. I have terrible problems with a local BCB radio station at 1150 KHZ that causes intermodulation distortion and/or blow-by detection at and below 200 KHz in addition to a terrible noise source at 78-120 KHz. Therefore, I built another cascade low pass/high pass filter and placed it in the same chassis as my regular NDB low pass/high pass combination filter for use when tuning LWBC and perhaps for when listening to frequencies less than 200 KHz.



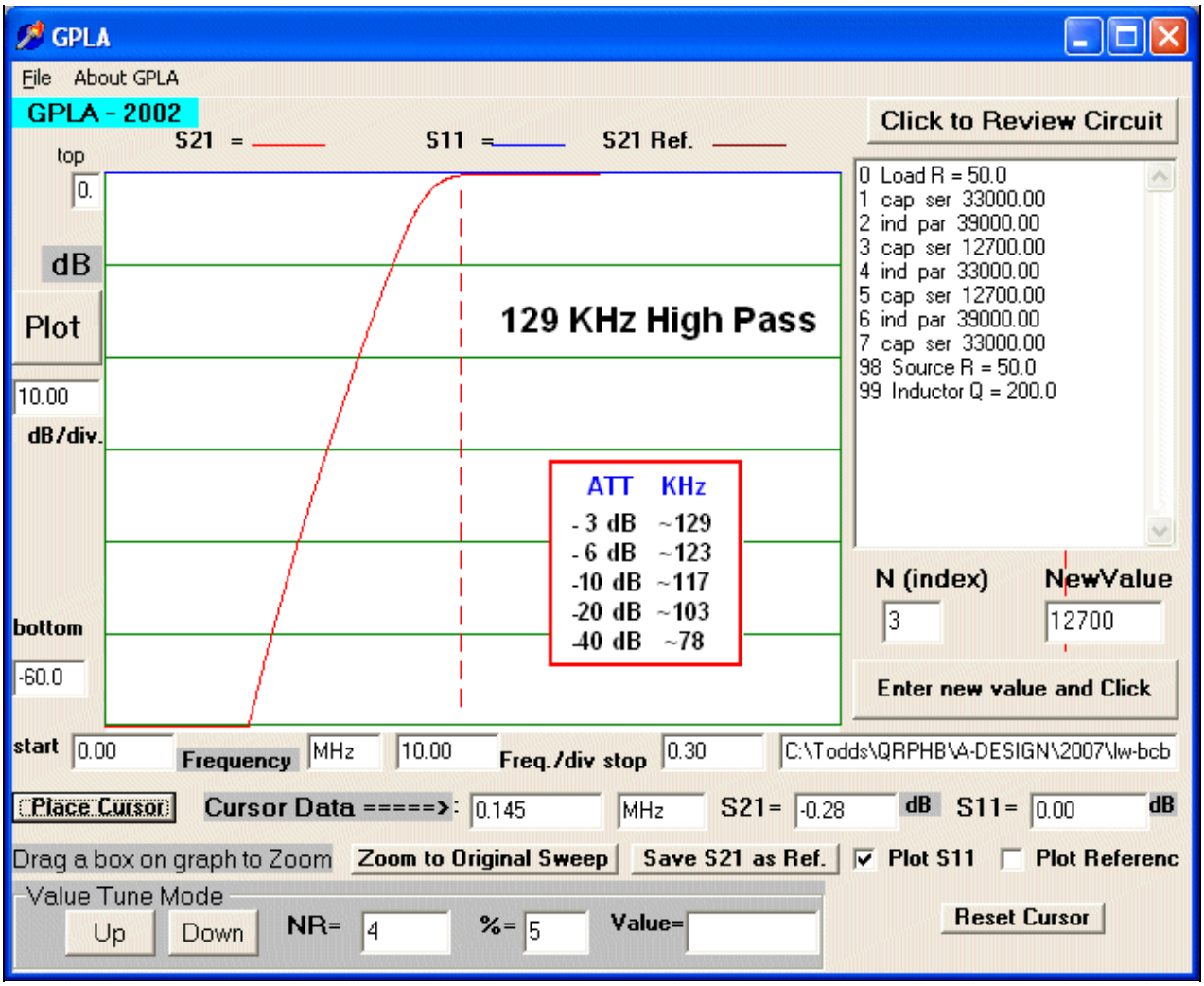
Above is. the schematic of the 322 KHz low pass filter. In the photograph above, you can see a 50 ohm pad at the input that was used only during testing. This filter offers a calculated attenuation of ~ 98 dB to my 1150 KHz interfering station. In reality it is not possible to achieve this level of attenuation, however, there is no detectable 1150 KHz signal interference with the filter in line which makes me happy. Click here to listen to the dramatic difference with regard to interference this filter makes at my QTH with my receiver tuned to 199 KHz. The receiver is set for wide band AM detection; first without the LWBC filter and then with the filter switched in. When the filter is switched in, the BCB interference disappears and a Canadian NDB (UAB @ 200 KHz can faintly be heard along with our cat meowing in the background. It is not possible to listen to LF without aggressive low pass/high pass filtering at my QTH.



Above is the GPLA simulation of the LWBC low pass filter.

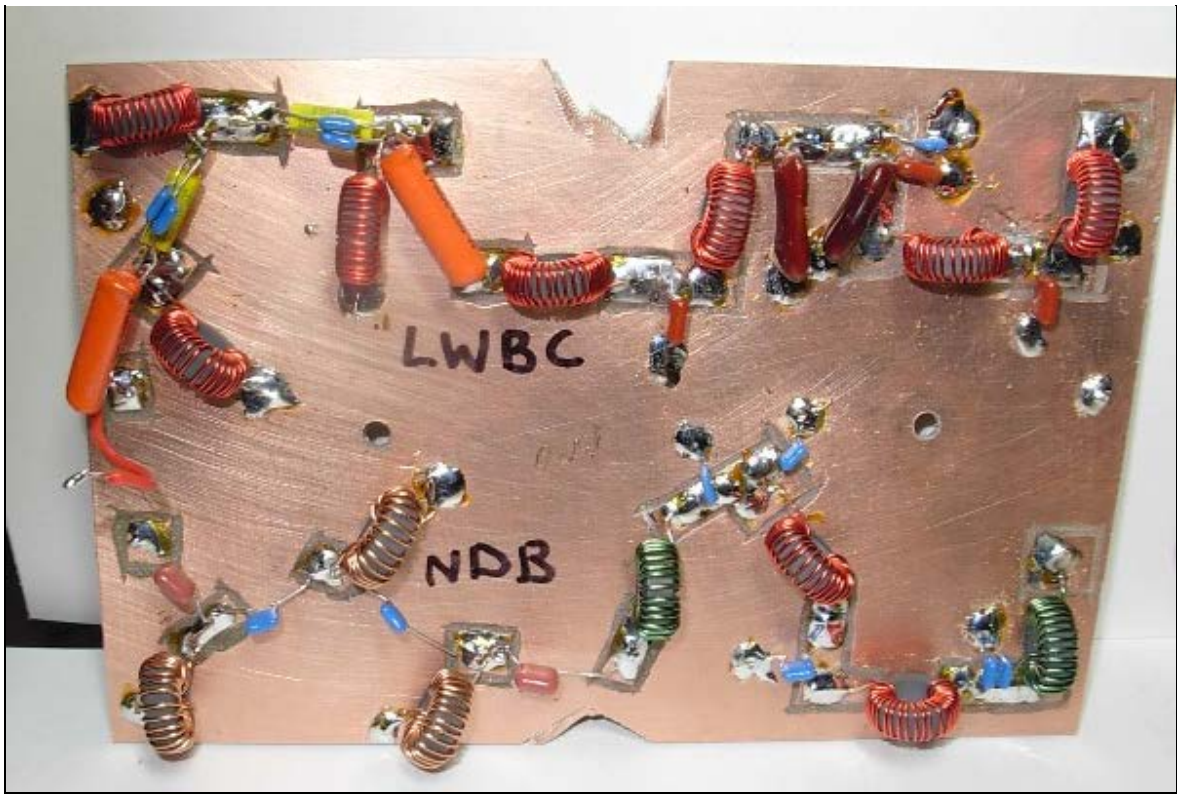


Above is the schematic of the 129 KHz LWBC high pass filter. In either of the 2 filters, capacitor values can be obtained using 1 or 2 standard value capacitors in parallel. The cutoff frequencies of both filters were chosen to allow using practical component values.



Above is the GPLA simulation of the LWBC high pass filter. The high pass filter might not be needed at your QTH. My LWBC filter has the low pass filter before the high pass filter. I.e. they are in series or cascaded.

Dual NDB and LWBC Filters



For use in my radio shack, I built LWBC and NDB filters inside 1 chassis with separate inputs and outputs. Some photos of this project are shown directly above and below. The NDB filter is the 526 KHz low pass filter in series with the 157 KHz high pass filter. The LWBC filter is the 322 KHz low pass filter in series with the 129 KHz high pass filter. High Q caps were used and the inductors were wound with either 22 or 24 gauge wire to obtain a relatively high unloaded Q. The large Hammond project case allowed reasonable spacing of the inductors and a nice long input to output layout.

Beacon XJ @326 KHz

Above is NDB XJ in Fort St. John, BC. Photo by VE7KPB in August 2008.

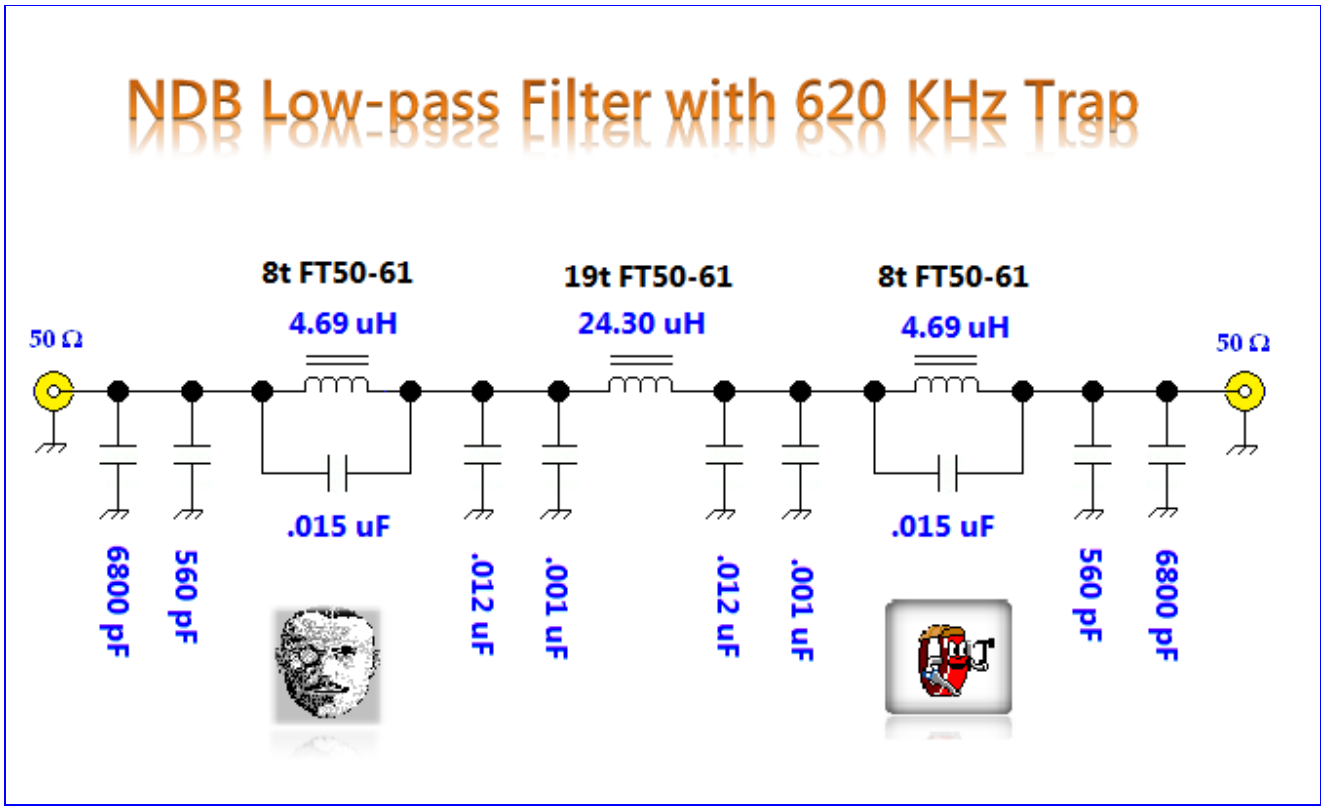
QRP — Posdata: NDB Low-pass Filter with Trap

An email from Rick, NU7Z spawned this 2012 addition.

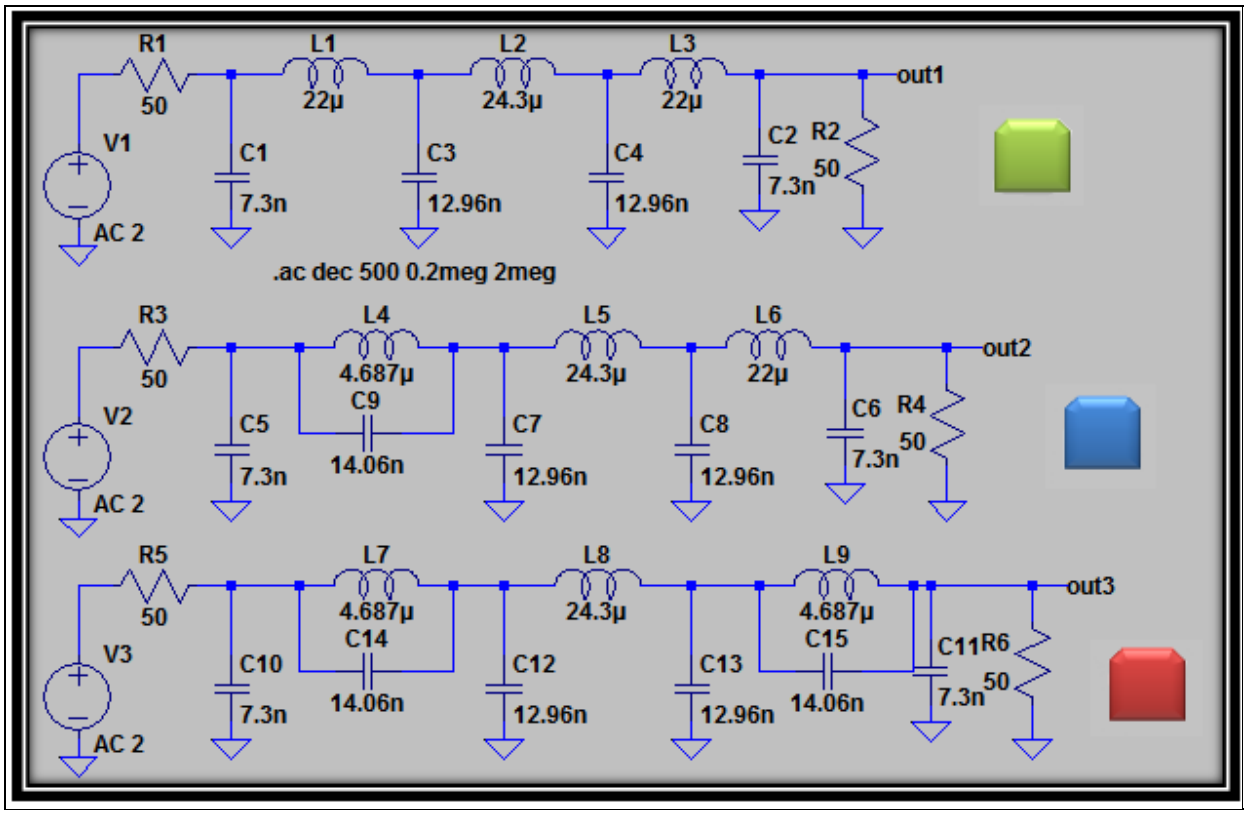
Depending on their design, typical NDB low-pass filters provide less than 20 dB attenuation at 620 - 630 KHz, and if you hear a strong

station on this frequency — good luck!

Rick sought a filter with a trap at ~ 620 KHz — after mulling around, we encountered design problems with a trap frequency so close to the low-pass cut-off frequency and later asked Wes, W7ZOI if he might help design our filter.



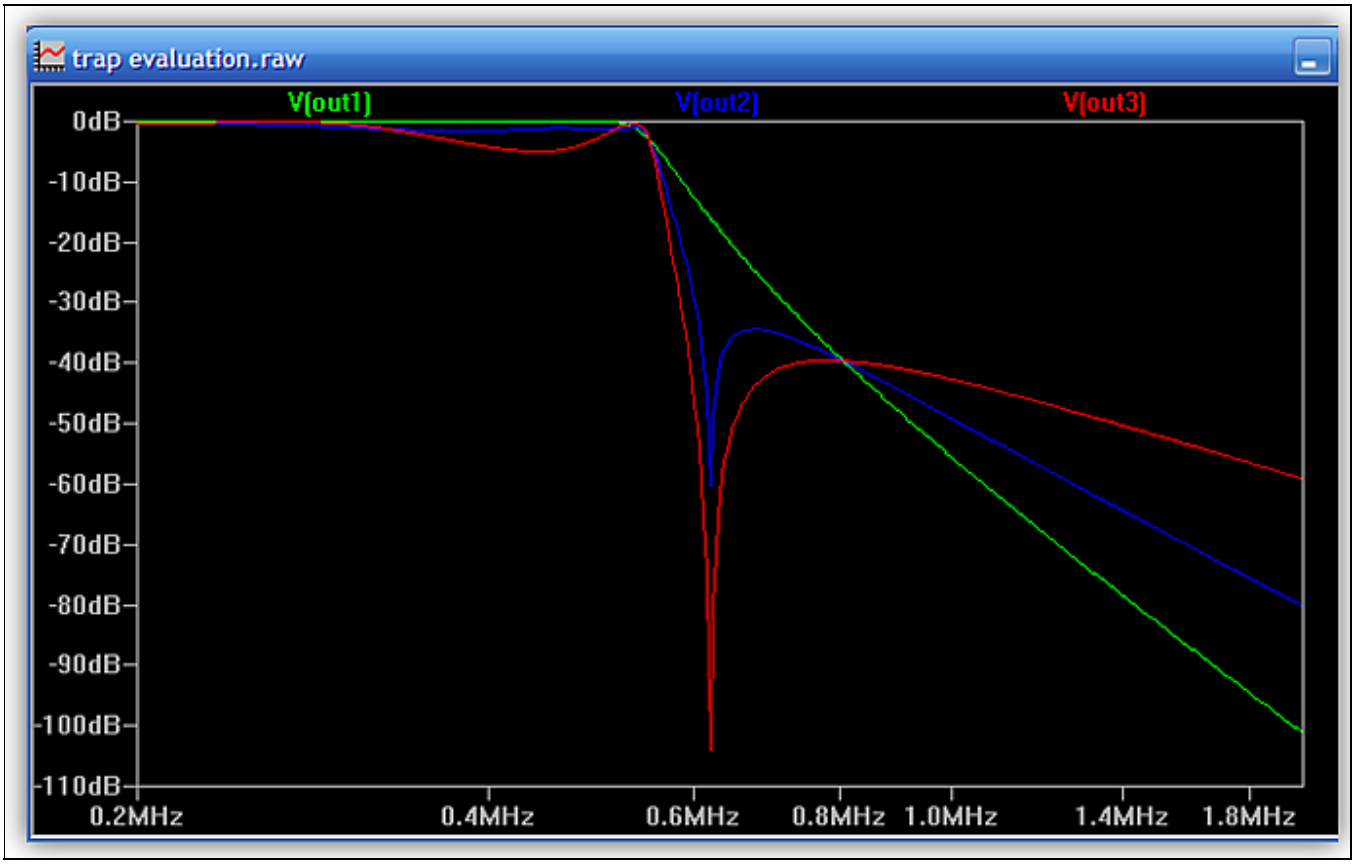
Above — My version of the Wes, W7ZOI designed NDB low-pass filter with a trap. **Red** filter below



Above — A 7th order, 0.1 dB ripple **Chebyshev low-pass filter** with a 550 KHz cut-off filter evolved to include **1 trap**, and then **2 traps** at 620 KHz.

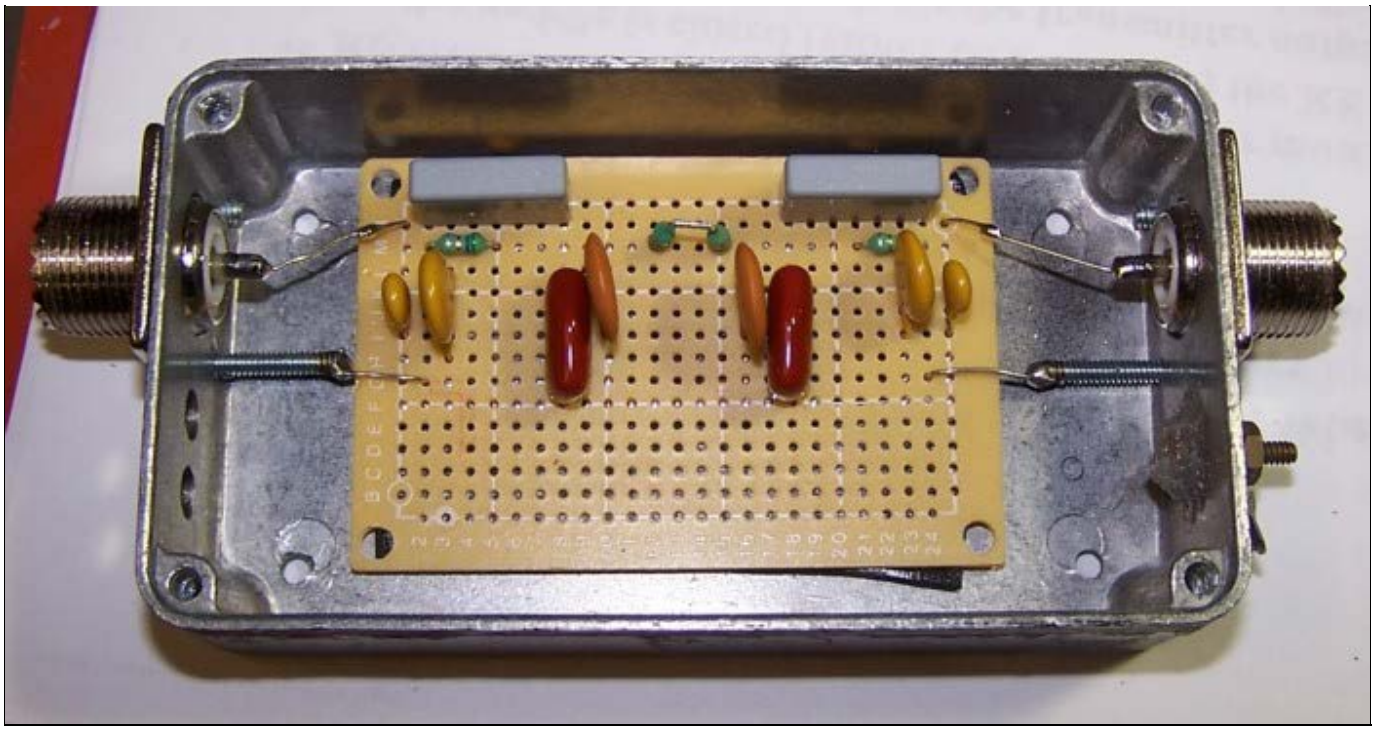
We learned that in simple situations, you may modify the elements of a low-pass filter so that the usual inductor is replaced by a parallel trap. See Wes' work in EMRFD Chapter 3; in particular, Figure 3.10.

Wes wrote he's employed this technique successfully before — for example, to add harmonic suppression to a simple output network for a QRP transmitter, although he hadn't added traps to higher order filters like the 1 we wanted. Click for a file containing the math contributed by Wes, W7ZOI.



Above — The SPICE analysis of the 3 color-coded filters above. This design excludes the impact of finite L and C and the unloaded Q that could significantly affect function since the trap frequency is close to the low-pass cutoff frequency. These factors usually worsen the insertion loss near cutoff, but since we're using this filter in a noisy RF environment, filter misperformance should be tolerable.

In the future, Wes recommended designing an elliptical low-pass filter with software such as that distributed by AADE.



Above — A version of the filter built by Rick, NU7Z using epoxy-coated inductors for the L's. The insertion loss with these inductors = ~ 5 dB, although he runs a 40 dB receive preamp and can accommodate such losses.

Despite employing a loop receiving antenna, he could not listen around 500 KHz due to a loud, local broadcast station at 630 KHz. Inserting this filter reduced this 630 KHz signal from 40 dB over S-9 down to S-1 on his receiver S-meter.

Fantastic! Big thanks (большое спасибо) to Wes, W7ZOI.

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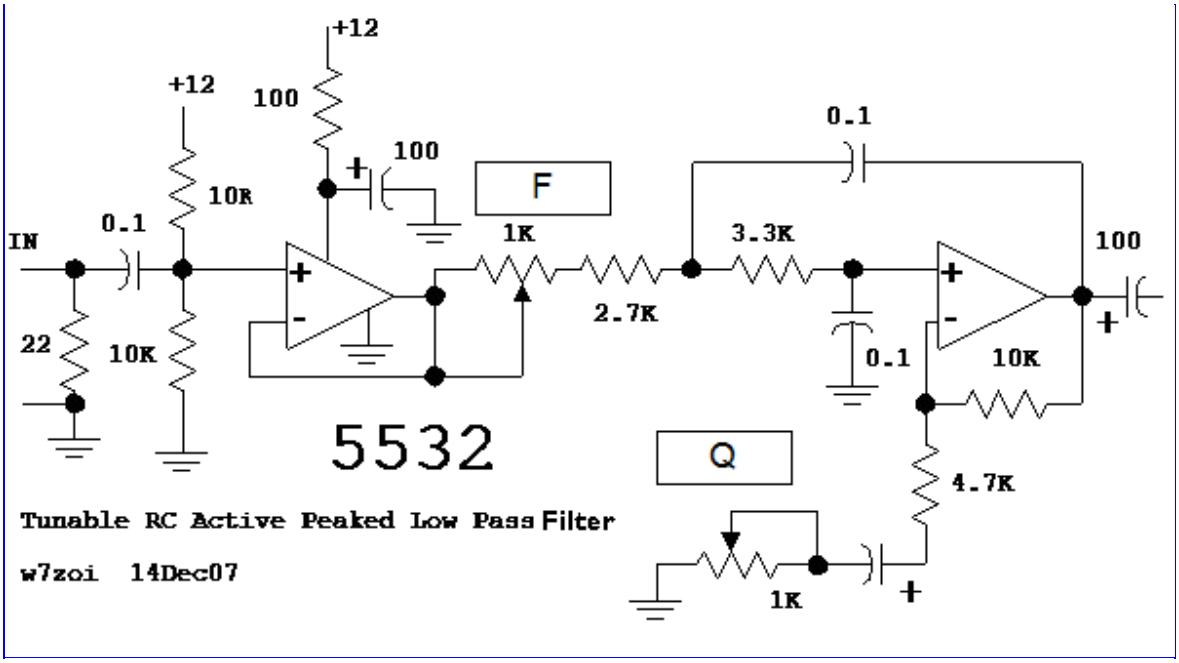
440 Hertz Peaked Low Pass Audio Filter Experiments

Introduction

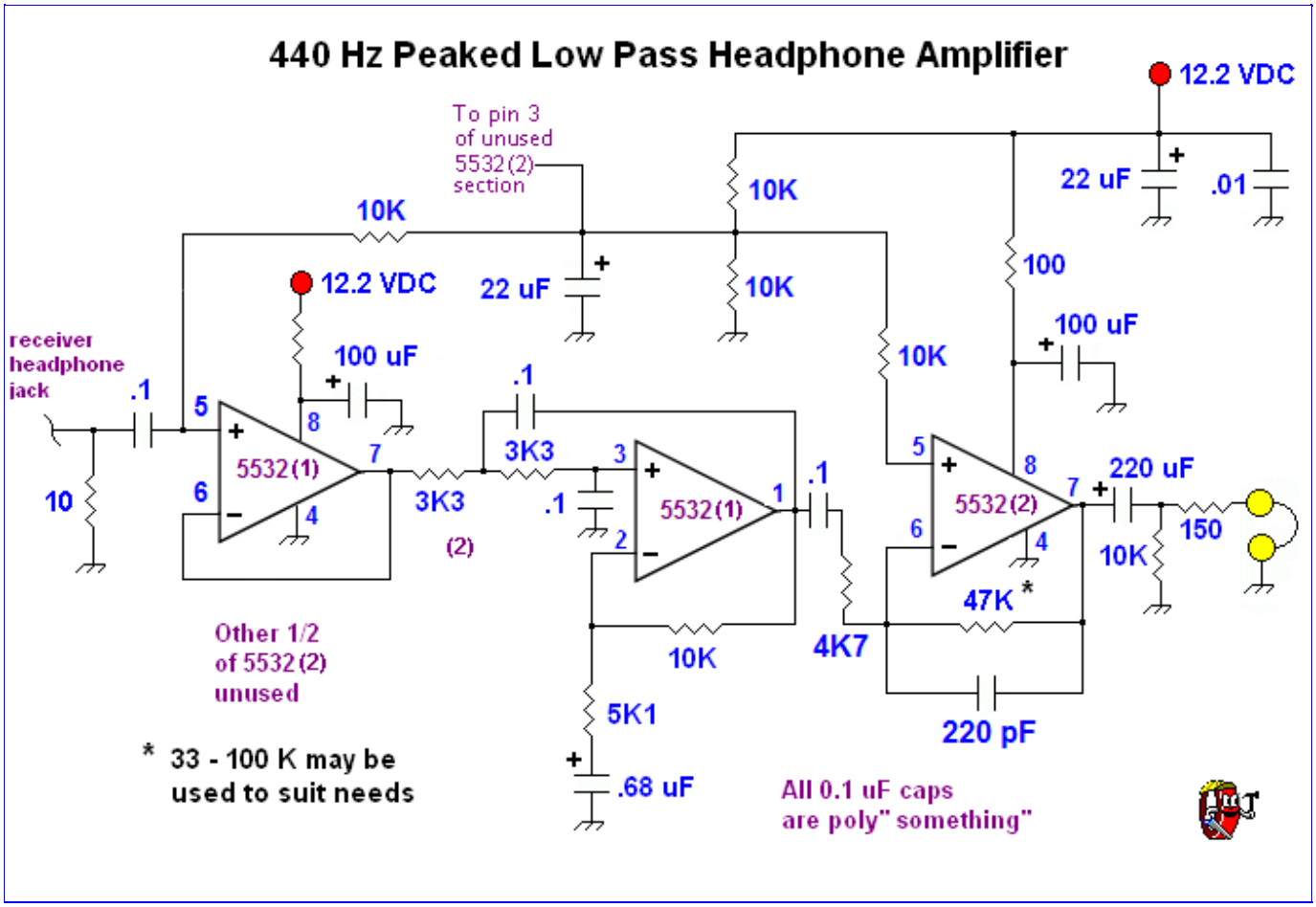
At audio frequencies, low pass filtering can go a long way to improving CW and beacon reception. It has a been a long time since I built one and therefore decided to experiment with some designs using the 5532 op amp. EMRFD has a great section on RC active audio filters starting at Chapter 3.5 and this is where I began. After some experimentation, I remembered a peaked low pass filter designed and published by Wes, W7ZOI in the 1970's. This filter became very popular in Russia after publication in a 1971 Russian Amateur Radio Journal. I asked Wes if he might design another low pass filter peaked for 440 Hertz, which is my favorite CW beat note frequency. This filter is intended for use as an out board headphone jack device for the Icom R75 or other receiver.

Base Project Schematic



Above is the filter designed by Wes, W7ZOI. It has two 1K pots you can tweak; a subtle frequency control and a Q adjustment. It is theoretically possible to adjust the Q too strongly as to cause oscillation, although this did not happen with my bread board.

Schematic and Circuit Building Details



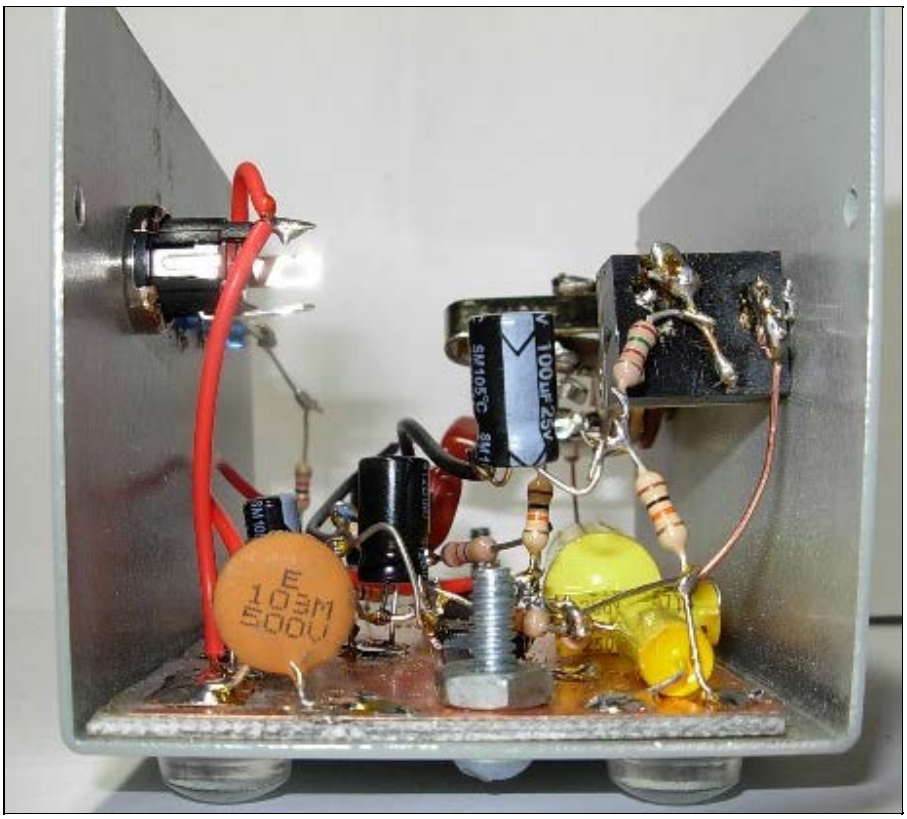
Above is the final filter design. My prototype filter had 2 variable pots and I disconnected and measured them after discovering my favorite setting

and then constructed the version shown above using fixed-value resistors instead of the potentiometers. If you like to tweak knobs, you might leave 1 or both of the pots in. The 0.68 uF cap can be raised as high as 4.7 uF if you have poly caps this large in capacitance value. although, it seems wasteful, I did not use one half of the second 5532 op amp. Feel free to add another pole of low pass filtering or something else to utilize this stage if you like. Pin 3 of the unused, second op amp 1/2 is connected to the 10K/10K voltage divider bias with a wire as indicated in the text.

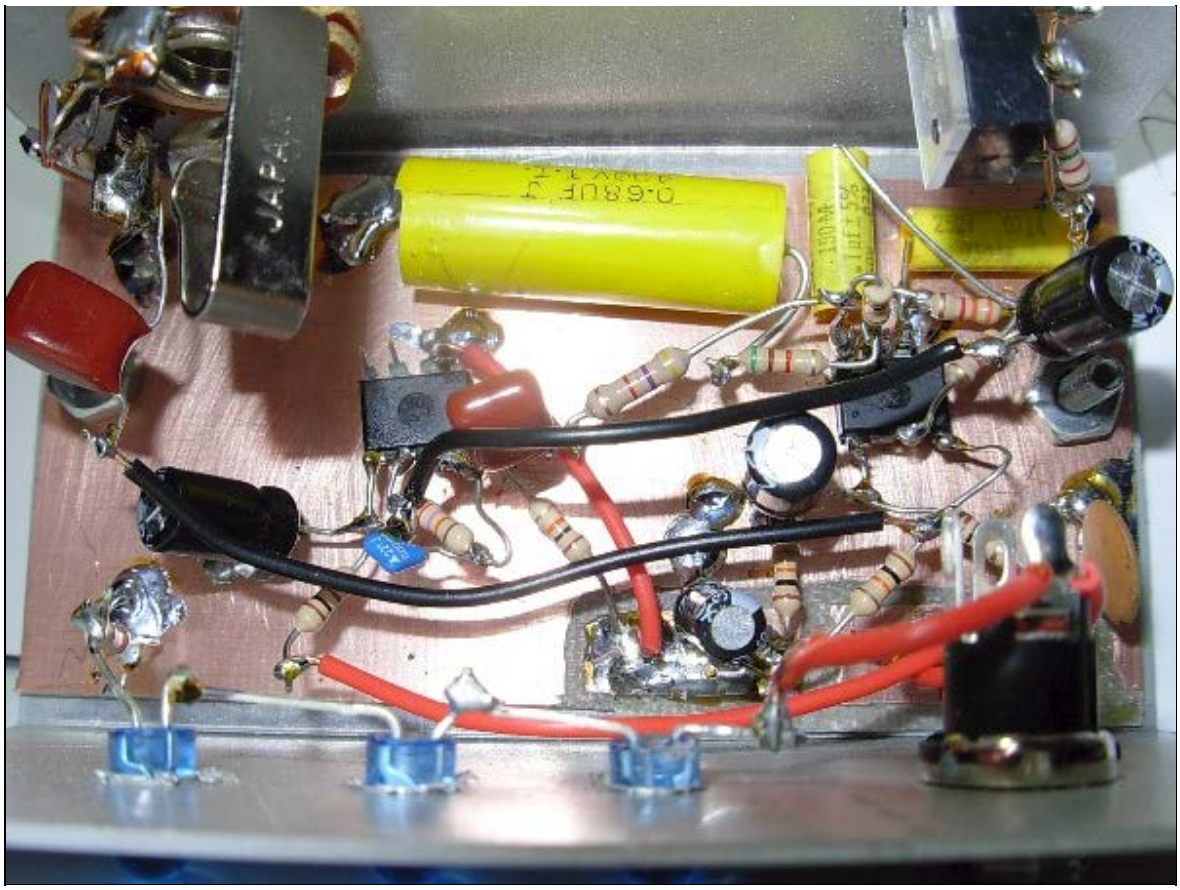
This filter sounds the best when the R75 volume control is minimally turned on as I suspect some of the wide band noise heard is from the ICOM AF chain. More importantly, If the R75 audio gain control is turned too high, the filter will be overdriven and sound distorted This is especially true when using the 2.4 KHz wide SSB filter on the R75. The 10 ohm filter input resistor attenuates the receiver output and makes it more difficult to overdrive the audio filter. The 500 Hz filter at the 9 MHz Receiver IF has quite a bit of of loss and with this filter switched in, it is difficult to overdrive the audio filter. For best results, an audio filter should be placed just after the first AF preamp stage, however, using the headphone jack is the only option available for adding AF filtering in most commercial receivers I have used.
The second op amp stage is used to increase the headphone volume and the 47K feedback resistor can be adjusted to suit your needs. It is really important to experiment with the component values which will match your receiver and the IF filters and antenna you have. For example the input shunt resistor may be increased from 10 to 18-22 ohms if you always use a narrow IF filter during CW and beacon listening or received signals are low in volume. This is an experimenter's circuit, not a finished project. Overall, this circuit has low output volume and is really gentle on the ears in terms of noise and amplitude.

Project Breadboard and Samples

Shown to the right is a side photo of the experimental project. The big yellow Mallory polyester caps were used as I did not have any other desirable AF filtering caps in my parts stock. The day after, I built this filter, my parts order (including a big selection of polyester film audio caps) from Digi-Key arrived, however this is Murphy's law. These Mallorys are good quality capacitors- just a little large!



Low pass filters can really help reduce noise during reception. Two example audio files follow. These were heard in the NDB band using a 2.4 KHz SSB IF filter.
290 YYF
312 UNT
The occasional scratchy noises are me moving as I held the headphones around the microphone.



Above is the filter photographed from above. The 3 blue LEDs are used to light the lower row of buttons on the R75 as it is difficult to see them with a low level of light in the radio shack. No hum is heard with this filter. When you build AF amps or filters with the 5532, after soldering pin 4 to the copper ground plane, start out by connecting the components associated with pins 6 and 7 and then 1 and 2. I suggest this as placing components between adjacent pins is often the most difficult part of building when using ugly construction with op amps.
If by accident you make an unintentional solder bridge between 2 adjacent pins on an op amp, simply heat the 2 bridged pins up and gently drive a small screw driver between the pins. This should remove your unwanted solder bridge. The 5532 op amp is quiet and relatively inexpensive. In EMRFD, there are countless examples of how one can use them in a variety of applications.

Additional Information and Photographs

Above is a photo of the audio filter in action on the NDB band. The blue LED reflection can be seen on the receiver. For serious NDB and CW pile up work, narrow band pass audio filters are generally required. This simple audio filter experiment might be useful as a spring board for your own AF filter experiments and to learn the filter requirements of your own particular receiver.

Above is the filter photographed from the rear. The DC power cord has a built in RFI filter. For homebuilt projects, DC power cords can be obtain by cutting the power cord off old unwanted or broken "wall wart" power supplies. This provides you with a nice cord with a built-in plug. I collect old AC "wall wart" transformers for this purpose. This filter is powered by the main 12 volt DC supply on my radio bench.

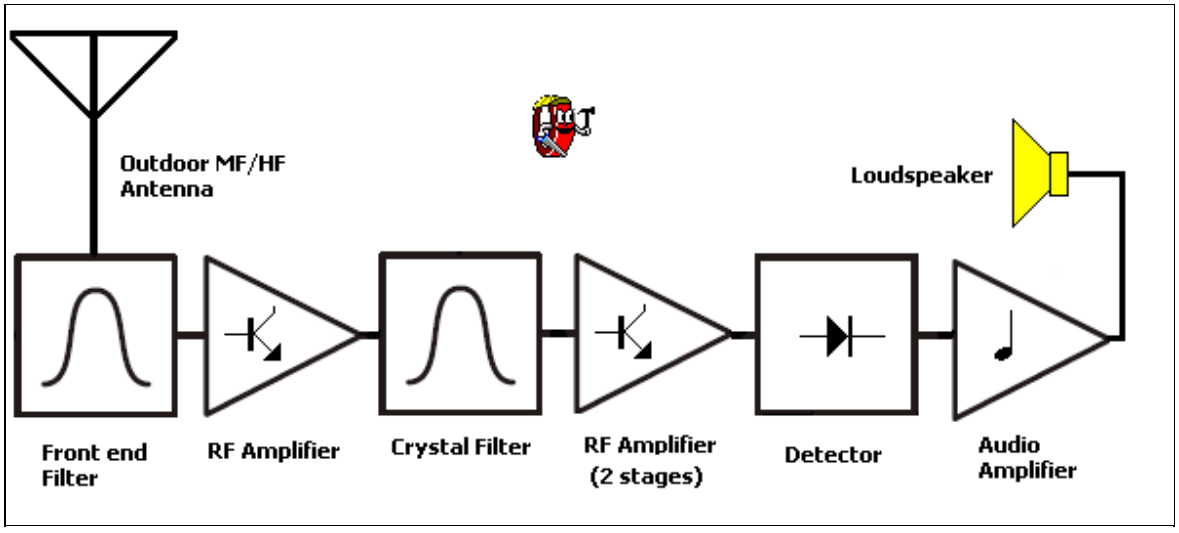
Cascode Hybrid-Based WWV Receiver for 5 MHz

Introduction

This was my favorite project of 2007. When I web published the original TRF WWV receiver for 10 MHz in 2006, there were many complaints that I used hard-to-find dual-gate MOSFETs and also that the AF stage lacked the popcorn factor that this web site has become strongly associated with. In this experimental project, these 2 concerns are addressed.

The cascode JFET and BJT amplifier stage used in this receiver is based upon the amplifer described in the Hybrid Cascode IF Amplifier article which was published in QST for December 2007 and designed by W7ZOI and WA7MLH. This amplifier topology has many advantages including high gain + low noise, that it can function well at DC voltages less than 12 VDC and that the noise figure does not degrade when the BJT bias (and stage gain) is lowered during AGC action. Please read the QST article and also refer to the W7ZOI web site for more details on the IF amplifier and the cascode hybrid topology.

Receiver Block Diagram



The receiver block diagram is shown above. To hear digitally recorded examples of this receiver, click here , here or here. No attempt was made

to make these files sound better than they really are- there is signal fading, room noise etc. A electret microphone was placed near the loud speaker to record these audio samples. *Note I am now compressing audio files in the mp3 format to allow listening by those who use Linux as their operating system.* **A supplemental web page** to this main web page is linked here

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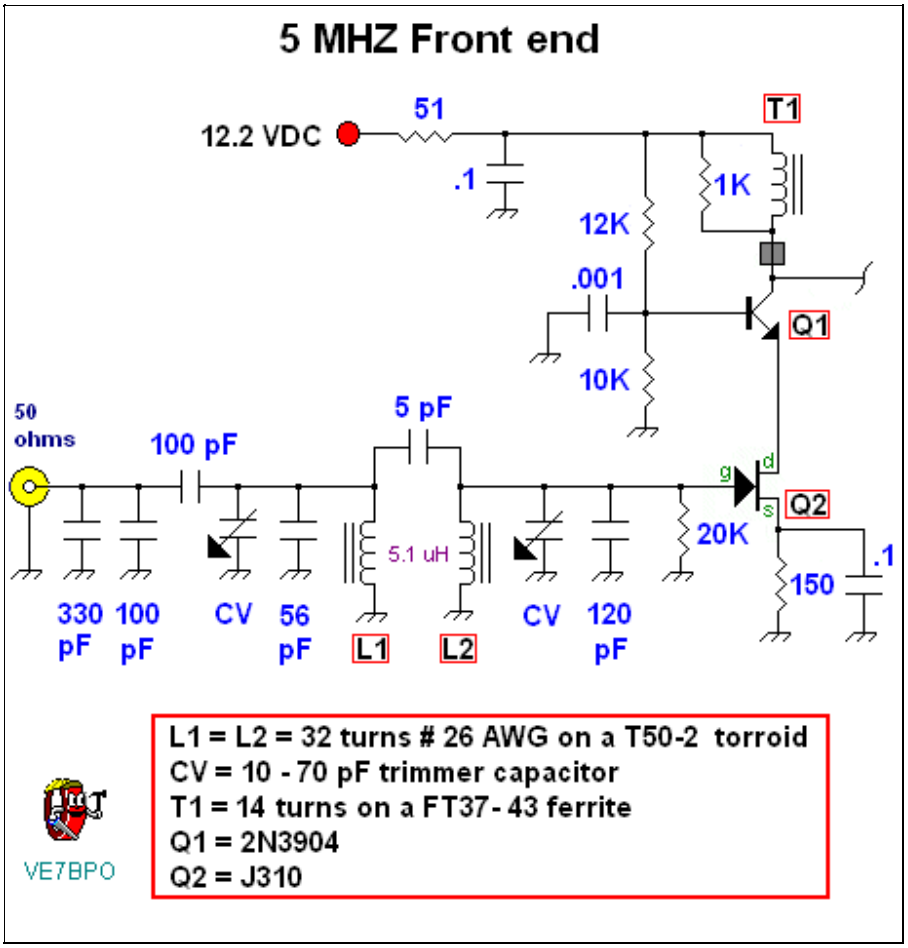
Receiver Front End: Band Pass Filter and First RF Amplifier



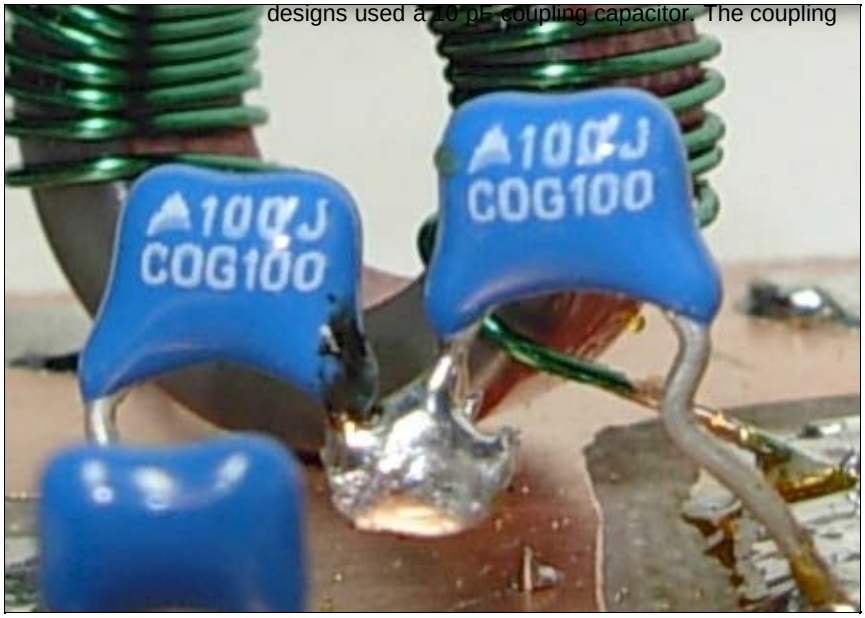
This receiver is meant to interface with a standard 50 ohm feed line. Testing was performed using my MF/HF antenna.
I built 3 separate bread boards of this receiver and tried varying the number of RF amps, using different detectors (as well as different detector followers) and eventually built and tested this basic receiver design for 5, 6 and 10 MHz. With respect to using the cascode hybrid amp (and probably any other amplifier type) **in a TRF receiver**, I learned 3 things:

1. Do not operate the RF amps at maximum gain. I built some very powerful amps with a Q2 source resistor of 47 ohms and over 6 volts bias on Q1. While powerful, this amp broke into oscillation and also consumed much current (nearly 20 ma).

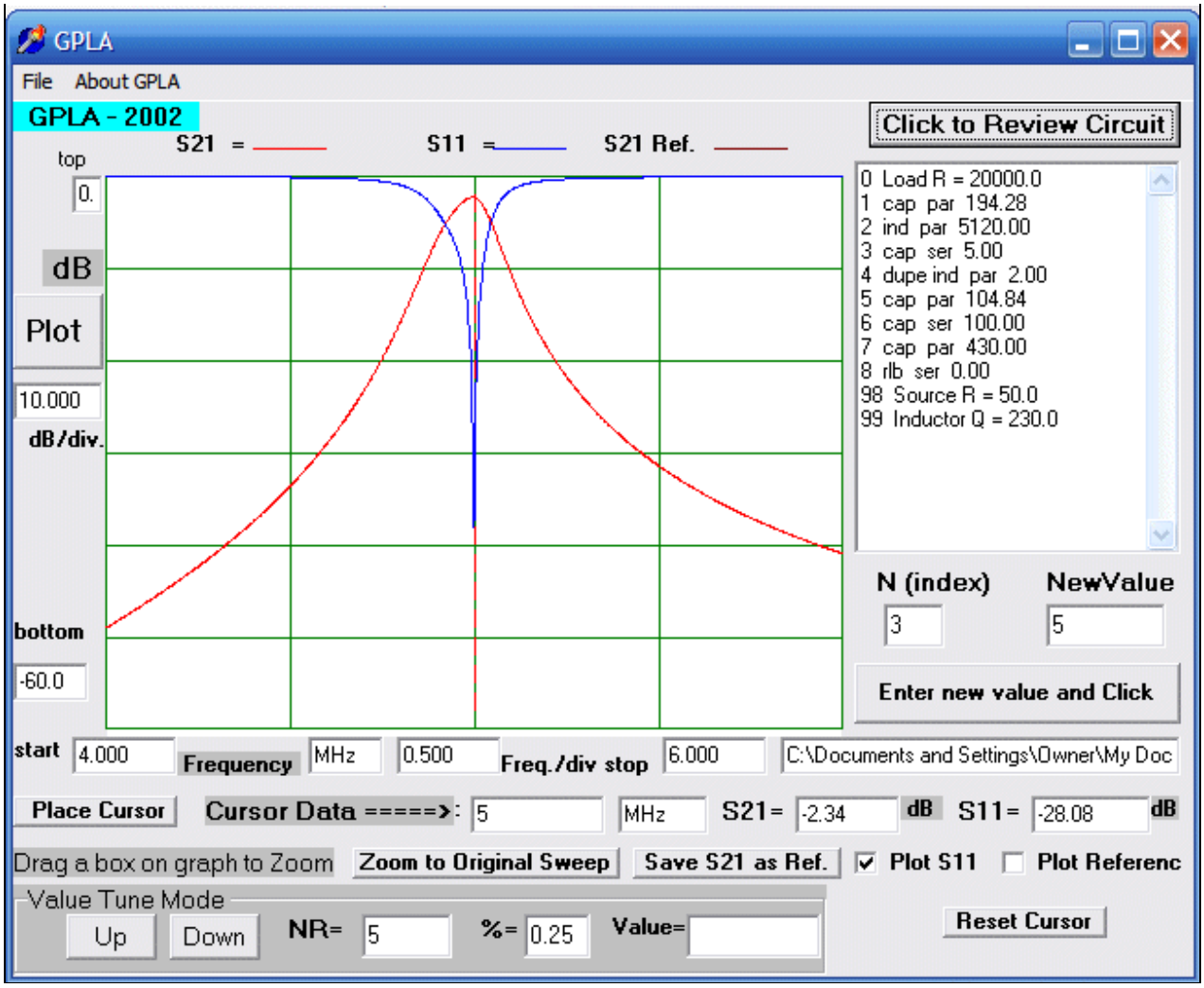
2. Keep the RF stages at least 2-3 cm apart to reduce the chance of parasitic oscillations.

3. Keep the input band pass filter at least 2 cm from the Q1/Q2 amp or you might encounter some unwanted oscillations.

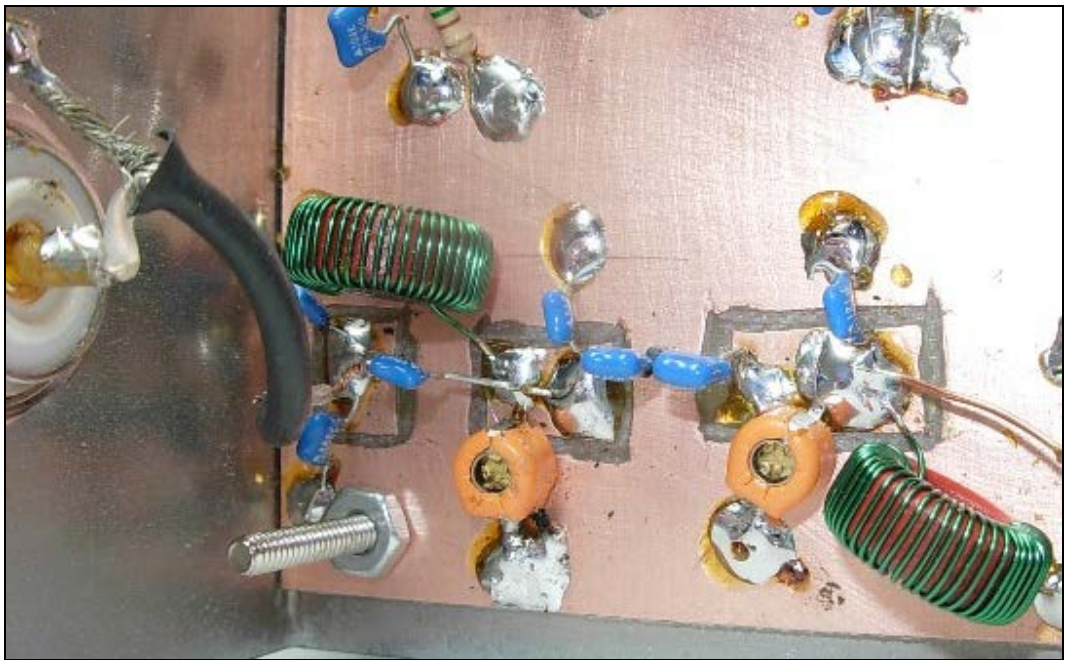
For the front end band pass filter, a reasonably narrow bandwidth was desired. When sweeping early filter designs using a signal generator and oscilloscope, a double humped response was noted. These filter designs used a 10 pF coupling capacitor. The coupling capacitor was then decreased to 5 pF. To obtain the required 5 pF, two 10 pF capacitors were placed in series as shown in the photograph directly to the right.



I struggled with this filter design because one end is terminated in the gate resistance of Q2 of the hybrid cascode amplifier and was not the standard 50 ohm impedance termination. My early filter designs suffered severe insertion loss or poor selectivity. I asked Wes, W7ZOI, for some instruction on solving my filter problems. I learned that this filter topology is referred to as a singly terminated, double tuned band pass filter. Wes designed the front end band pass filter for the 5 MHz receiver for us all to learn from and for this I am very grateful to him.



Above. A GPLA simulation of the singly terminated, double tuned filter designed by W7ZOI. A double tuned circuit is mandatory ahead of the WWV receiver as local BCB and other RF energy will be amplified by the first RF amp and may distort the WWV signal in the crystal filter or even might blow-by the crystal filter and be detected and heard in the speaker.

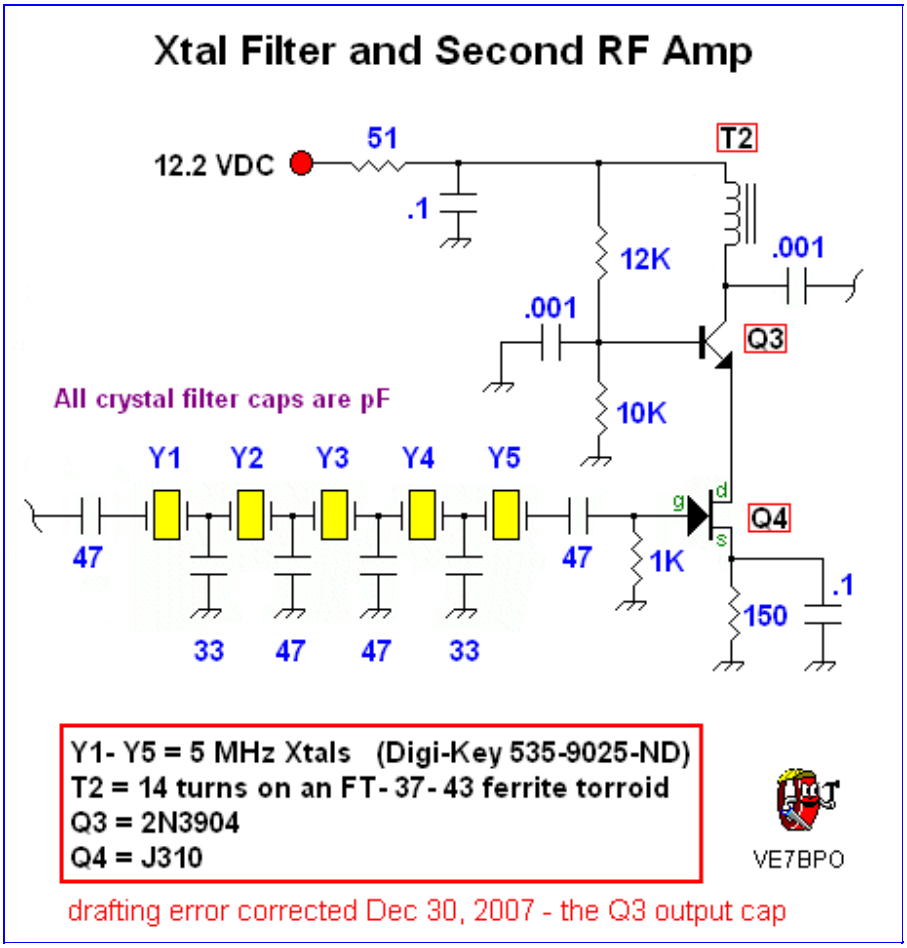


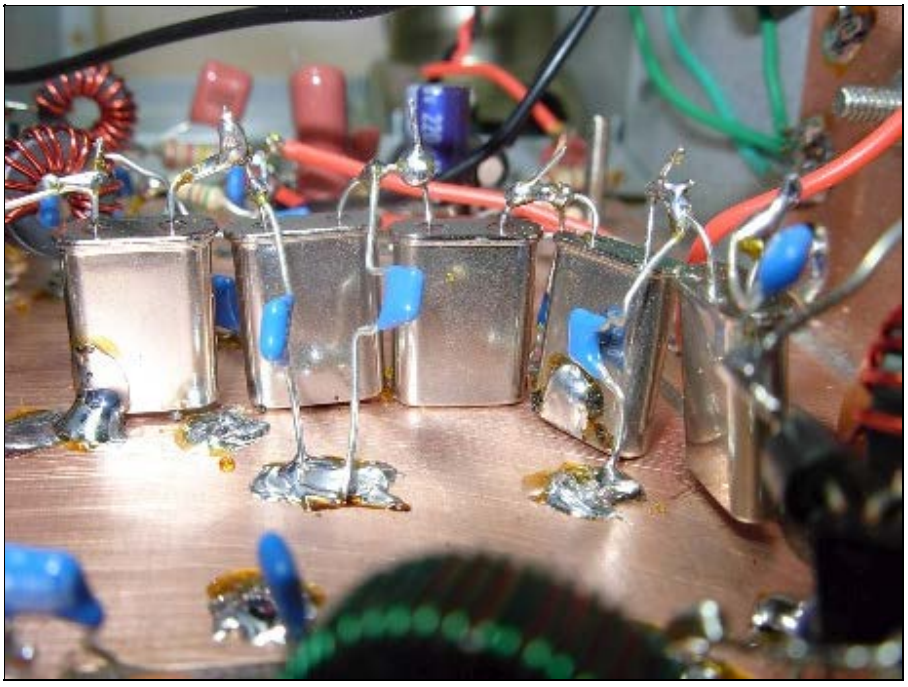
Directly above is a close up photograph of the input filter bread board. Filter tuning was done by ear (and screwdriver!) Simply tune the trimmer capacitors for the loudest audible WWV pulses in the speaker and you are set. If you can't locate a 20K gate resistor for Q2, a 22K resistor will work okay.

Crystal Filter and Second RF Amplifier Stage

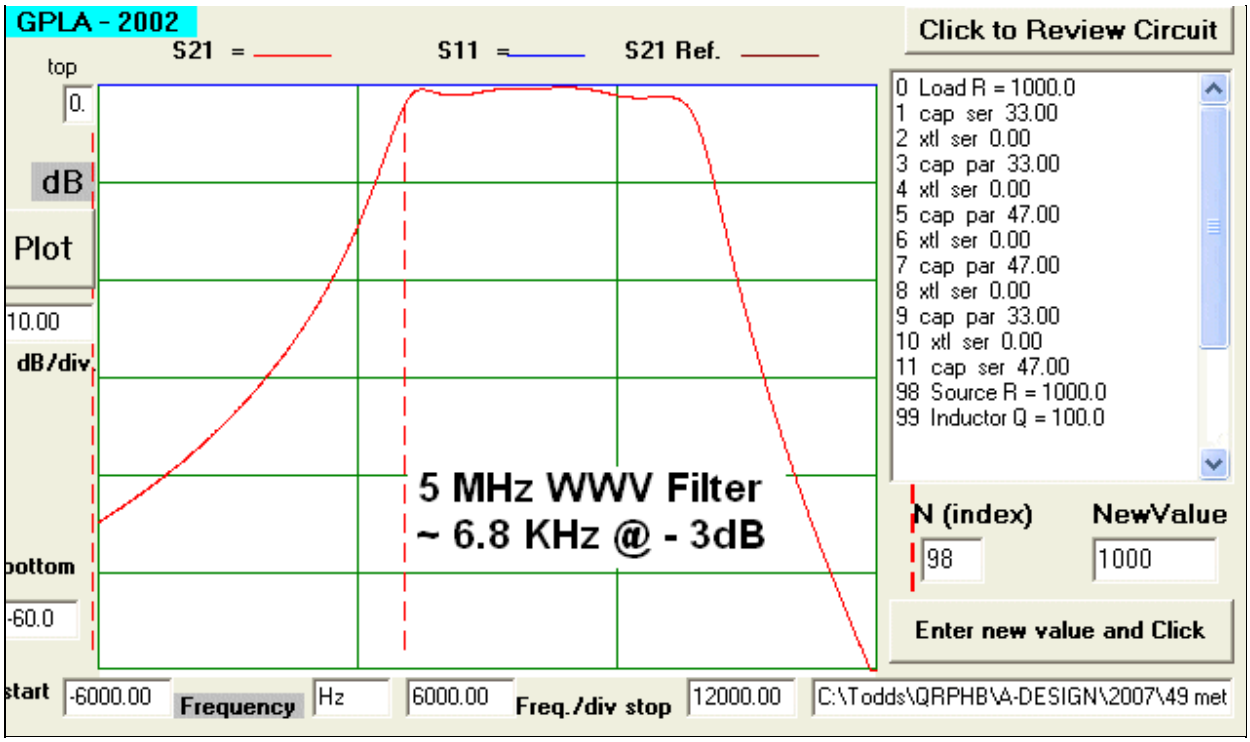
In the schematic to the right is the crystal filter and second RF amplifier. The input impedance of the crystal filter is established by the 1K shunt resistor across the output transformer on Q1. The output impedance of the crystal filter is set by the 1K gate resistor of Q4. A filter input/output Z of 1000 ohms gave the best overall shape and bandwidth during my testing.

Developing this filter was difficult. My first batch of junk box crystals had a low motional inductance and with the filter I built I could hear stations ~400 KHz below and/or above the filter center frequency in addition to WWV. After giving up in frustration for nearly 2 months, a batch of 10 crystals were ordered from Digi-Key. These were microprocessor crystals; ones with 18 pF load capacitance in a HC49/U holder. The new filter was tweaked and tested and now provides single signal reception of WWV. Your own results may vary according to your crystal parameters. The Digi-Key part number is provided for reference purposes only.





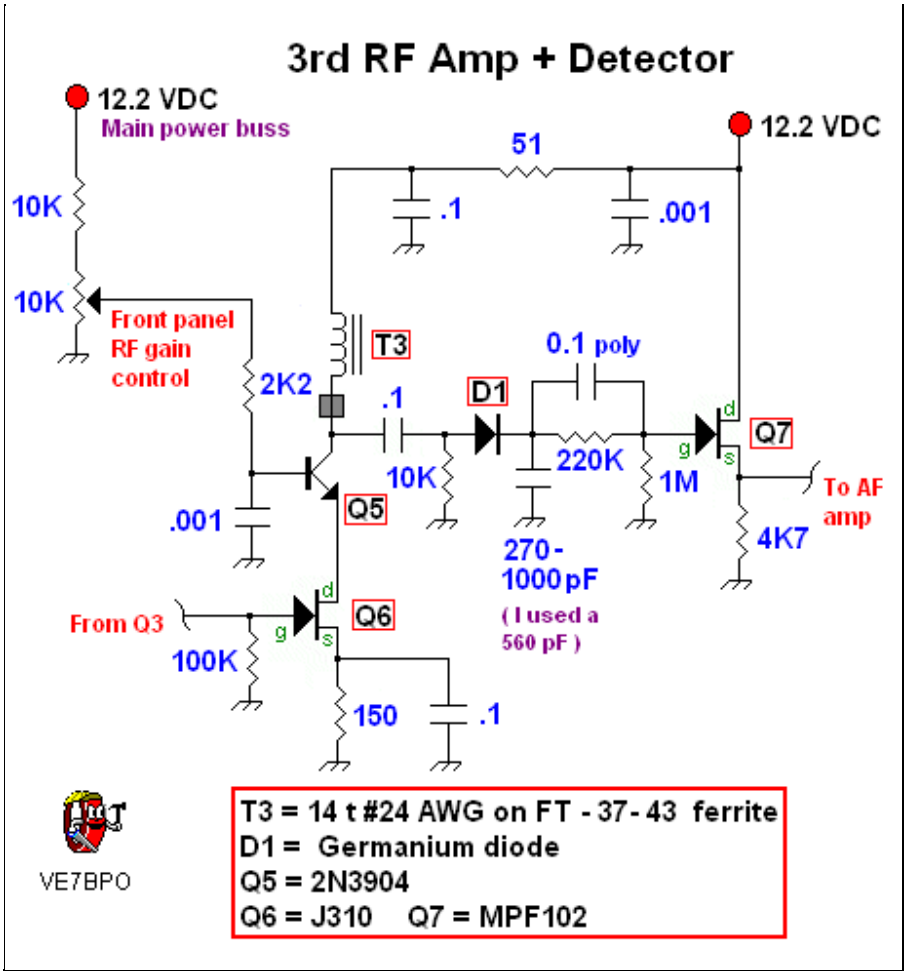
To the left is a close up photograph of the 5 MHz crystal filter. The crystals were turned upside down and the outer cases were directly soldered to the copper ground plane as you can easily see in the crystal to the left of the others. The rest of the crystals as well as one of the 47 pF tuning capacitors were soldered on the other side and solder points are hidden from view. The crystals were positioned to keep the output of Q1 away from the input of Q4. ***Stage lay out is very important in TRF receivers.*** I found stage layout to be far more important than keeping lead lengths short from my experimentation.



Directly above is the GPLA simulation of my crystal filter. The 5 MHz point is not centered exactly in the middle of the pass band, but a reasonable AM filter was built nonetheless. Crystal parameters, especially motional inductance and capacitance can make or break your filter. Motional inductance and capacitance describe the L and C values that make up the crystal's electrical LC model. Very large inductive and capacitive reactance values at the specified operating frequency give the crystal its extraordinarily high "quality factor" or "Q". For example, If the motional L is too low, your filter may not work as expected; providing single signal reception of WWV. The Lm was 0.02 and the Cp was 5 in the crystals which I used for my filter. In general, low Q crystals will give poor results. Oppositely, crystals with very high Q may give a lower then expected bandwidth and this may reduce AM receive fidelity. Experimentation is necessary.

Third RF Amplifier Stage and Detector

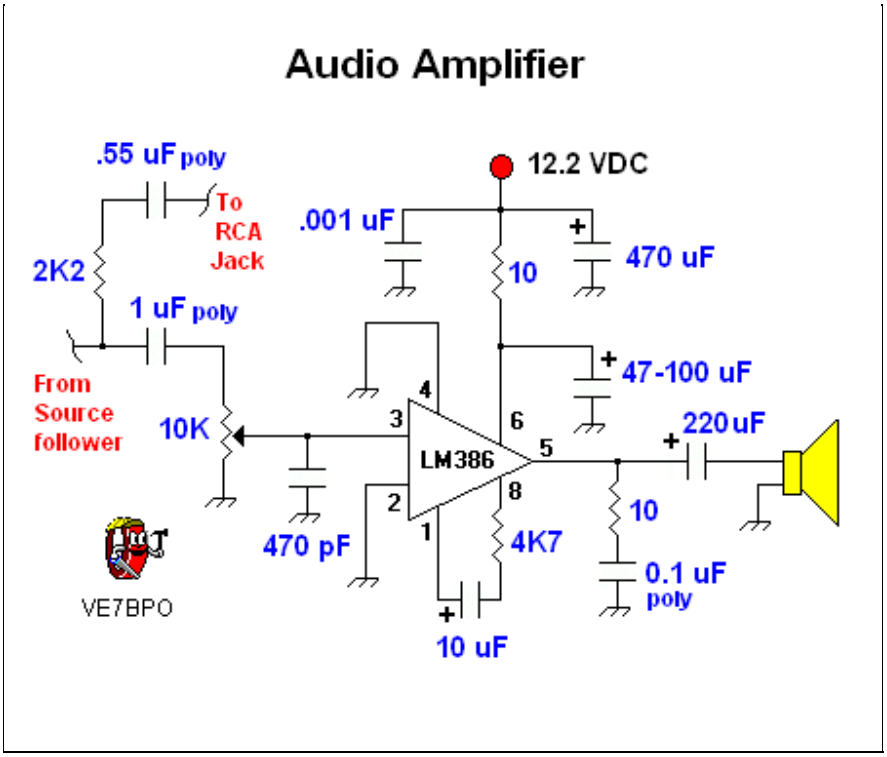
To the left is schematic of the final RF stage and the envelope detector. This RF stage has variable gain by means of a front-panel mounted 10K potentiometer which is used to vary the bias on Q5. The input Z of this stage is 100 K and is set by the Q6 gate resistor. The output of Q5 is AC coupled to a detector designed by Wes, W7ZOI. I performed considerable experimentation with basic diode detectors as well as detector source followers; some of which I sent to Wes for his consideration. He designed and emailed me back this simple, good sounding detector design which uses the gate voltage of Q7 to bias the germanium diode. Other types of diodes such as as hot carrier diodes will likely not have the output voltage of the Germanium type. Germanium diodes, when biased, had more noise and high frequency response in addition to higher output when compared to others I tried during my experiments. Diode detector guru, Felix, VK4FUQ advised me of an excellent diode he is now using called the BAT46. The audio samples of a local AM radio station using this diode and his other hi-fi lab equipment that he sent me are beyond fantastic.



The photograph on the right is a close up of the enveloped detector designed by W7ZOI. The germanium diode was purchased from The Source in Canada (Radio Shack in the USA). The blue, partially hidden shunt capacitor is a multi-layer ceramic 560 pF cap. The other capacitors are metalized, polyester film types. Ensure correct diode polarity.

Audio Stage

To the left is the schematic of the audio stage. The very "popcorn" LM386 AF chip is used to please the audience who complained about my AF stages not having enough popcorn factor. A 4K7 resistor was inserted between pins 1 and 8 to reduce the gain somewhat. Thus, the LM386 is still being operated in the high gain mode but won't hurt your ears with loud noise and distortion. The 470 pF cap on pin 3 may be changed or eliminated. It is a simple low pass filter.

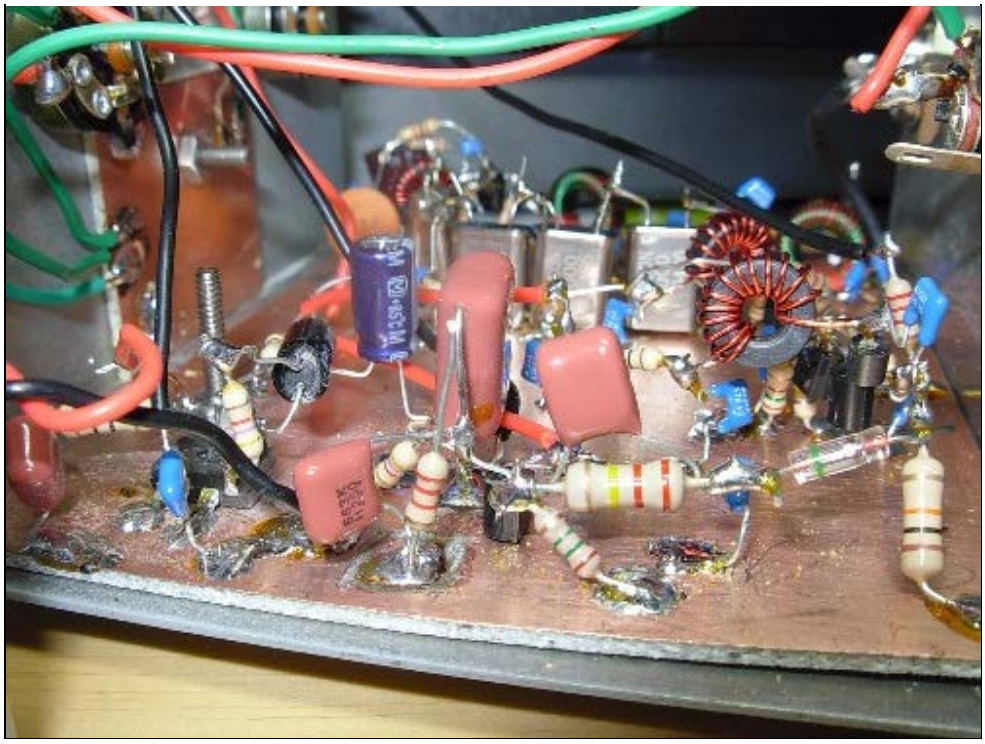


A secondary, audio output connects to a front panel mounted RCA phono jack. This allows me to use my lab grade (KK7B AF amp) and turn the audio off on the normal receiver AF amp.

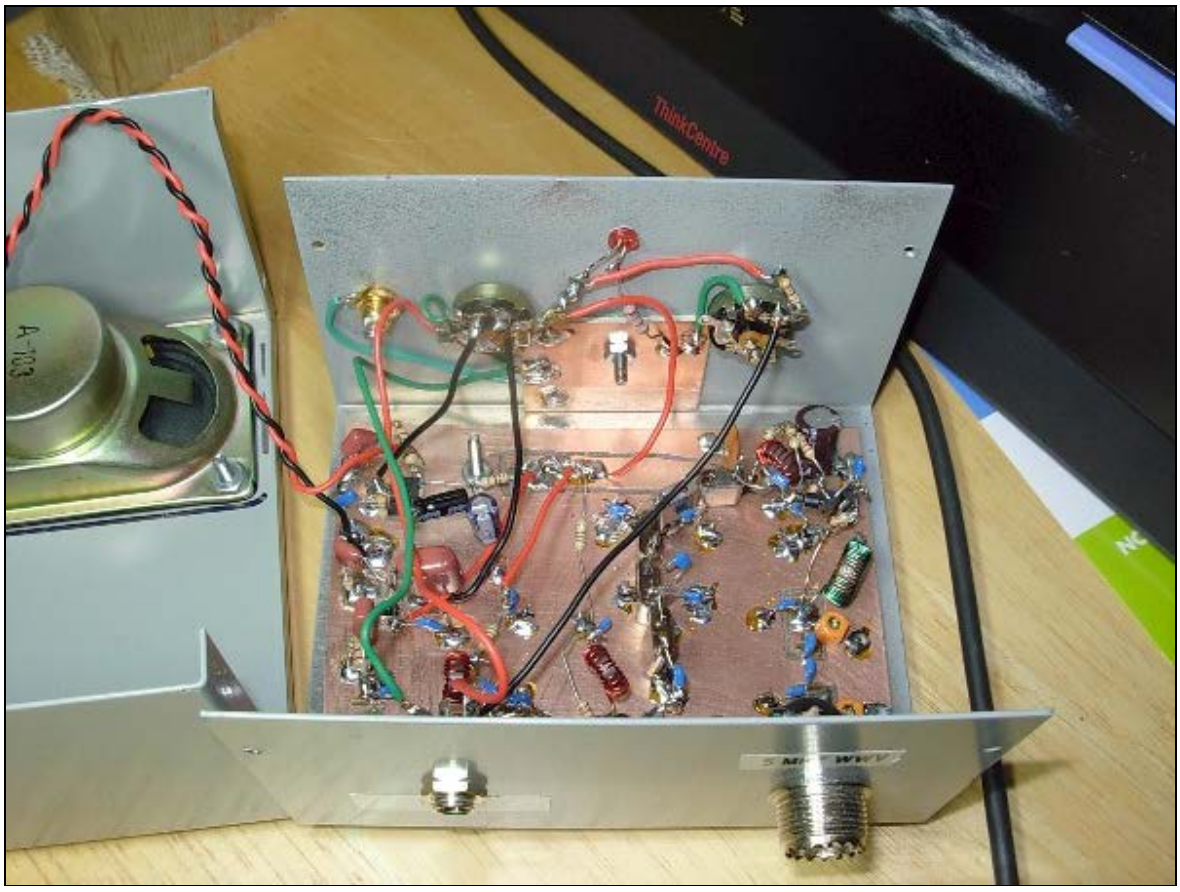
The photograph to the right is a close up of the LM386-based audio stage. This is where I started. After drilling the chassis, wiring the speaker, installing the chassis potentiometers, making the main power buss and LED indicator, the AF stage was built on the main board. The main board was then temporarily soldered in and tested.

(Some connections were made via alligator clips such as the speaker wires). When the AF amp worked as expected, the main board was removed from the receiver chassis and the net stage was built. Up next were the detector and source follower. After bread boarding these, again the main board was laid in the chassis, wired, tested and then removed when all was functioning well. To test the detector I touched my finger to the input and heard local BCB radio. Following this, RF Amp #3 was added to the main board and again the main board was temporarily wired up and tested by touching the input of Q6 with my finger and observing that a local broadcast radio station increased/decreased in amplitude when the RF gain control was turned up and down. DC voltages were also measured and considered from project start to finish.
Actually, all you need to do is connect a band pass filter such as this to Q6 and the components after and you will have a nice TRF BCB AM radio. Each successive stage was built and tested, so when the receiver was finished, I already knew that it worked. I cannot emphasize enough how important it is to build your receiver backwards and test each stage as you go. There is strong temptation to start at the antenna connection and work until you get to the speaker, but please consider doing the opposite.
The bare copper wire in the photograph is the positive connection point for the speaker wire. It was trimmed somewhat during final assembly to reduce the possibility of it shorting.

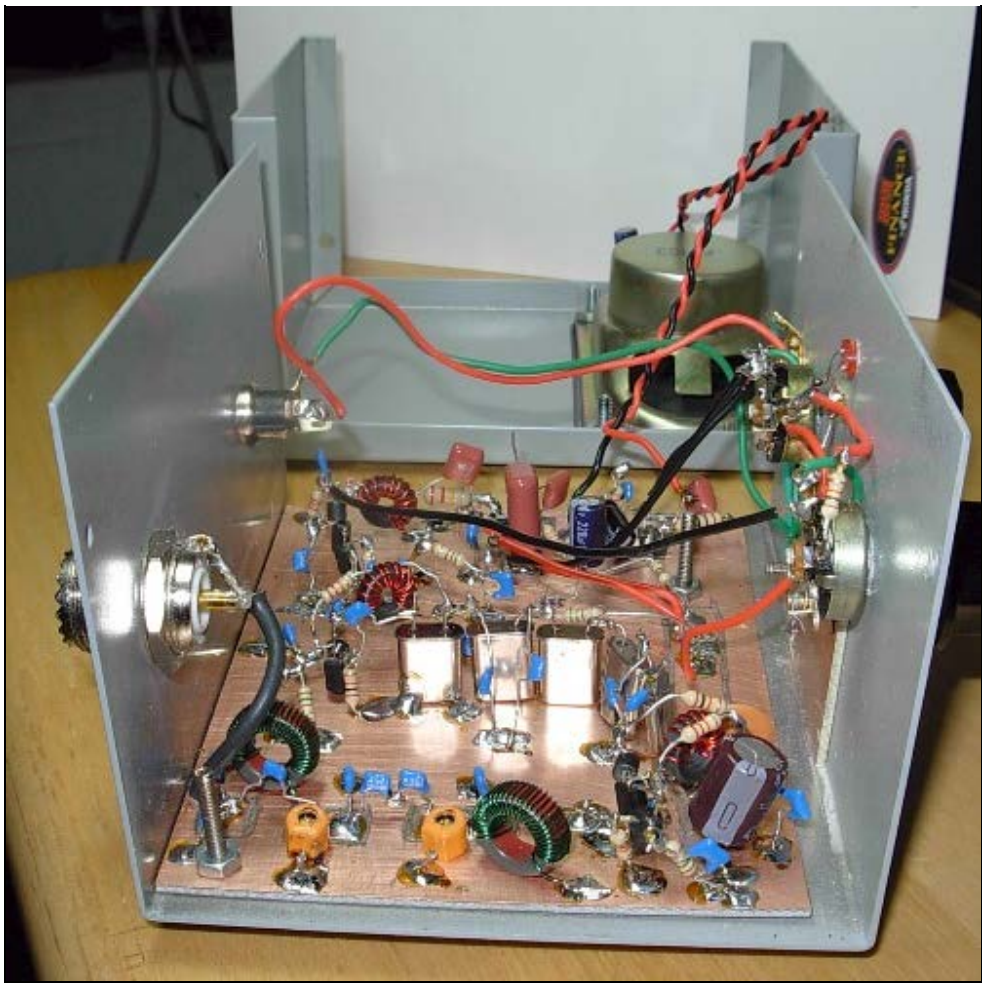
Miscellaneous Photographs



The photograph above shows some of the detail of the receiver main board from the right hand side which contains the detector, source follower and audio amp stages from right to left.



The photograph above shows a top view of the main chassis and also the chassis cover with the speaker bolted on and wired up.



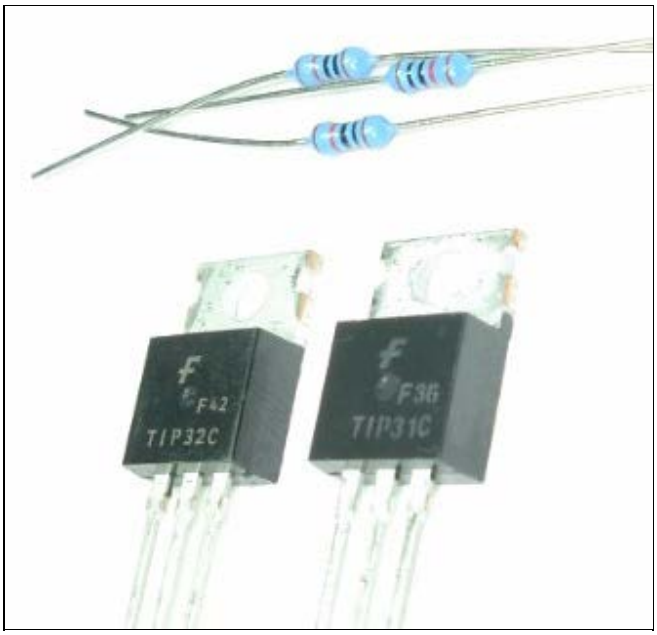
This wider angle photograph shows the main board from the left side. From left to right in the nearground are the SO-239 antenna connector, LC band pass filter and first RF amplifier.

The photograph above shows the speaker attached to the Hammond chassis top. Holes were drilled in the chassis lid with a drill press to allow the sound to pass through.

The photograph above shows the reverse view of the receiver chassis.

Complementary-Symmetry Amplifier Biasing Basics

Introduction

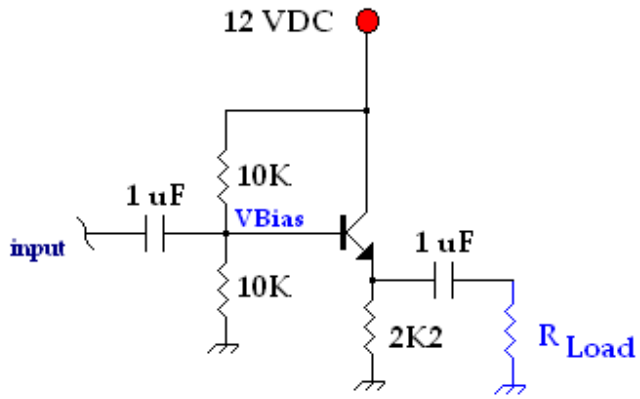


This page provides information concerning the biasing of Class-AB, complementary-symmetry audio amplifiers. These schematics should be considered theoretical, as design considerations such as thermal stability, negative feedback and component power ratings are minimized or excluded for sake of clarity.

The basic 2 transistor complementary-symmetry amplifier may be used as a simple low power AF amp or as a building block for a high powered stage such as a 50 watt guitar amplifier. It is important to understand how to properly bias your AF power amps to reduce distortion and to promote easy troubleshooting when problems arise. This web page describes the hows and whys of biasing in a progressive manner with minimal math.

Discussion

Figure 1



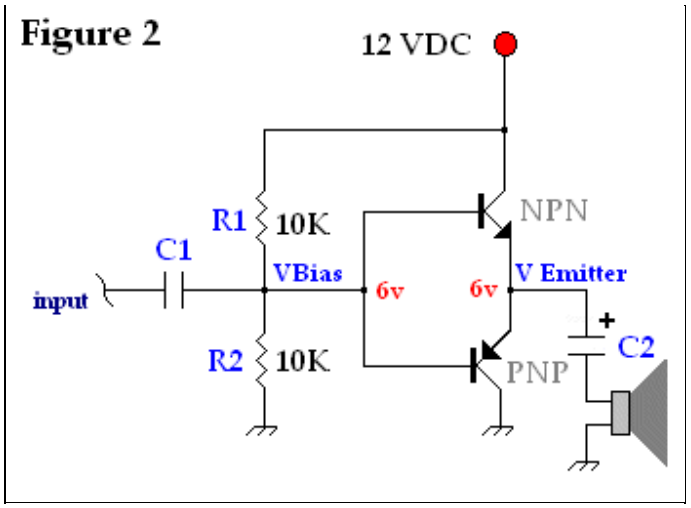
A review of the common collector amplifier (which is more commonly called the emitter follower) is a good place to start. We may refer to the complementary-symmetry transistor pair as complementary emitter followers since they are an NPN and PNP emitter follower connected in series. An emitter follower amp is shown in Figure 1. Its properties include:

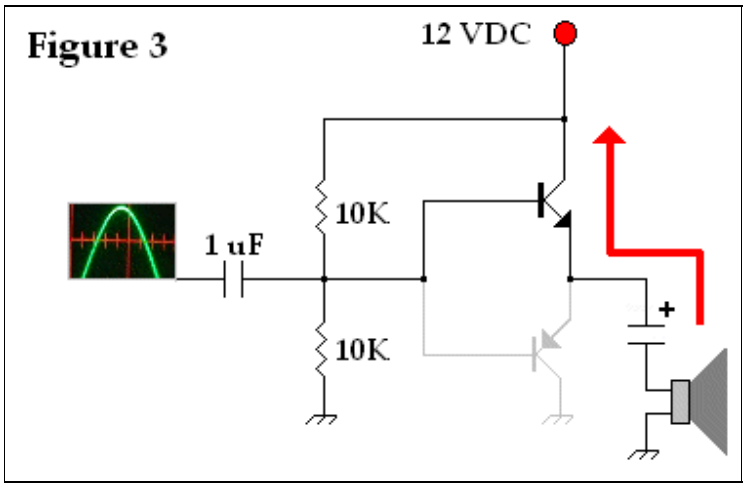
- ✓ input on the base - output on the emitter
- ✓ high input impedance and low output impedance
- ✓ a voltage gain of 1
- ✓ good current and power gain

In an appropriate configuration, these qualities are perfect for driving a low impedance load such as an 8 ohm speaker with large output currents that are not provided by our typical transistor or op amp voltage amplifier stages. In many cases, we bias the emitter follower with a voltage divider network comprised of 2 identical value resistors. In Figure 1, the voltage divider consists of a series pair of 10K resistors and thus VBias = 6 volts. These 10K bias resistors will be used throughout this web page as the circuits evolve.

In Figure 2 is a pair of complementary emitter followers which have their

bases biased with our now familiar series connected 10K bias network for a VBias of 6 volts. When the power is turned on, output capacitor C2 charges through the NPN transistor until it reaches about 6 volts *(theoretical value used to keep things simple)*. When the voltage at point V Emitter reaches 6 volts, the NPN transistor goes into cutoff because VBias voltage now equals the V emitter voltage. Recall that the NPN transistor base must be positive with respect to the emitter for current to flow. Both the NPN and the PNP transistor are in cutoff. This is the amplifier's quiescent state *(assuming no signal is applied to the input via C1)* and is called Class B bias.

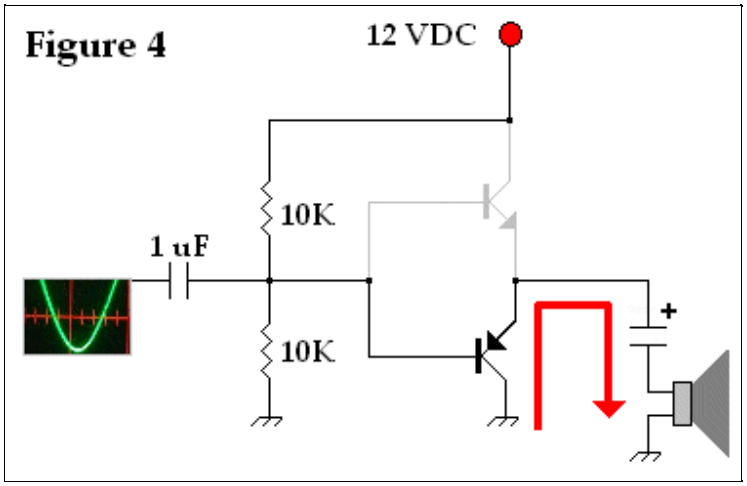


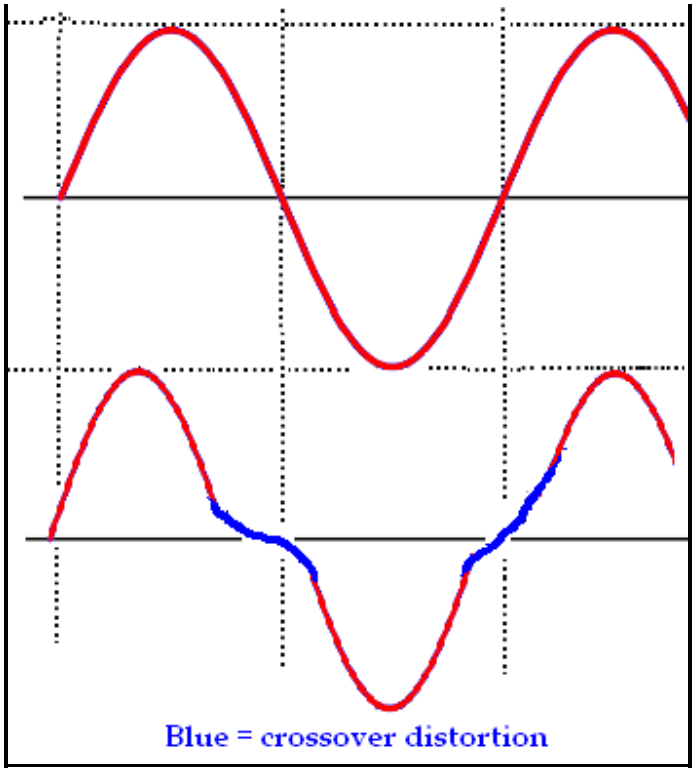


In Figure 3, a positive going signal is applied to the input capacitor. The NPN transistor becomes forward biased and turns ON. Current flows **through the NPN transistor** and charges capacitor C2 to a higher potential. The PNP transistor stays in cut off. The NPN transistor is an emitter follower connected to the speaker.

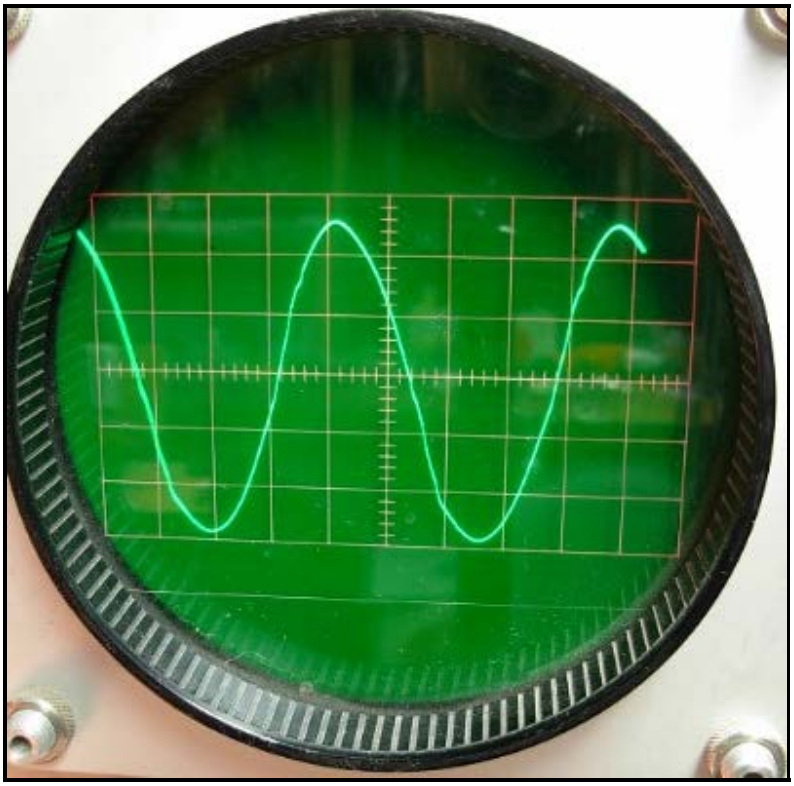
In Figure 4 a negative going signal the (negative half-cycle) is applied to the input. Q3 turns ON and discharges the output capacitor through the speaker as shown in **red.** The PNP transistor is an emitter follower connected to the speaker.

Thus the NPN and the PNP transistor conduct on alternate half cycles which causes AC current to flow through the speaker. The complementary emitter followers are said to be in push-pull operation.





The circuit of Figure 2 has a significant problem; output signal distortion. Silicon transistors such as the 2N3904 and 2N3906 will not conduct until their bases are forward biased by somewhere around 0.7 volts. For the NPN transistor, this means that it will not conduct until the input signal has gone positive by about 0.7 volts. Oppositely, the PNP transistor will remain in cut off until the input signal goes negative by approximately 0.7 volts. As a result, there is a dead zone during the point in time when one transistor cuts OFF and the other turns ON. Shown above is a normal sinusoidal AC waveform in red and another with the distorted waveform of Figure 2 in red and blue. This distortion is called crossover distortion because it occurs at the zero crossing point of the AC waveform. This introduces odd-order harmonics into the output signal. Such is the drawback of the Class B amplifier.



The above photograph shows crossover distortion in an under-biased power amp.

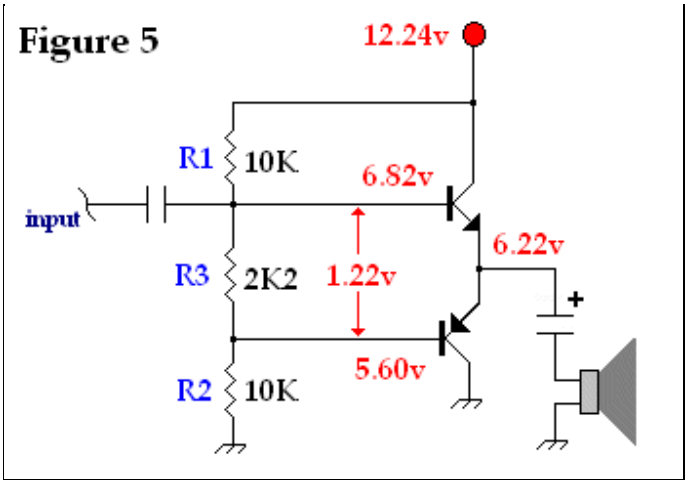


Figure 5 shows the principle technique used to reduce crossover distortion; both transistors are (slightly) forward biased almost to conduction in their quiescent state. As a result, any amplitude of positive or negative going signal will bias the appropriate transistor into conduction. An easy way to achieve this biasing is by adding 1 resistor to our 10K voltage divider network. In Figure 5 is a circuit I built, measured and listened to. R3, a 2K2 ohm resistor was placed in between R1 and R2, our usual 10K bias resistors. As a result , both transistors are forward biased.
That is: the base of the PNP transistor is negative with respect to its emitter and the the base of the NPN transistor is positive with respect to its emitter. As a rule of thumb, you need to drop at least 1 volt across R3. I chose a 2K2 resistor and it worked fine in my particular amp. The Figure 5 biasing topology is rarely used as it puts a series resistance on the PNP input among other problems; however, it exemplifies the basic principles of biasing our complimentary pair. With the forward bias on the transistor pair we now are in Class AB. The output capacitor serves to block the quiescent DC current from flowing through the speaker.

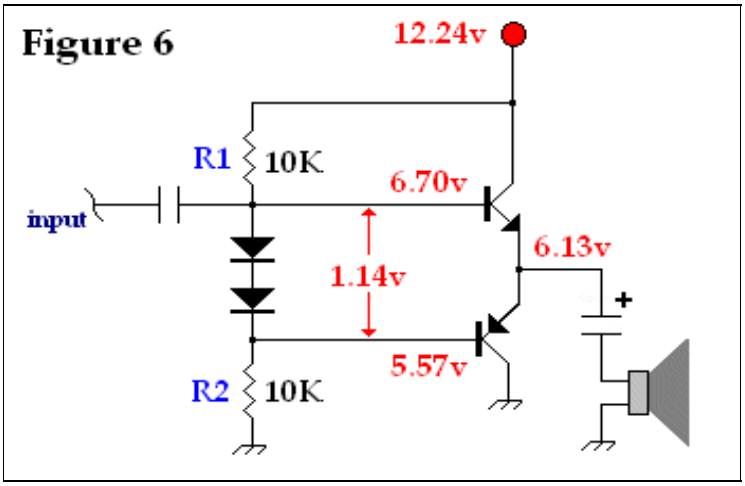


Figure 6 illustrates an improved biasing method over that of Figure 5 by using a pair of silicon diodes. You see this circuit used a lot by hobbyists. The voltage divider consists of 2 resistors and the 2 diodes. The 2 series connected diodes are connected in parallel to the NPN and PNP transistor base-emitter junctions which serves to keep the transistors turned on slightly. The net effect of the diode pair is the same as R3 in Figure 5. The voltage drop per diode was measured at 0.57 volts. The AC resistance of these 2 forward biased diodes is non-significant. There is major problem with the diode/resistor voltage divider; no way to adjust the diodes forward voltage drop. If each diode's forward threshold voltage is unequal to the base-emitter junction voltage of each transistor, either not enough forward bias is applied, or the 2 transistors may be turned on too much reducing efficiency and possibly cause excessive heating. Additionally, the pair of diodes lack the ability to provide temperature compensation when the transistors get hot.

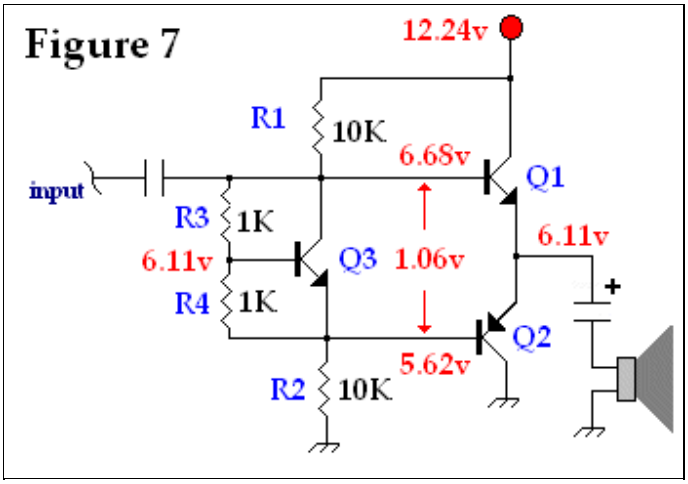
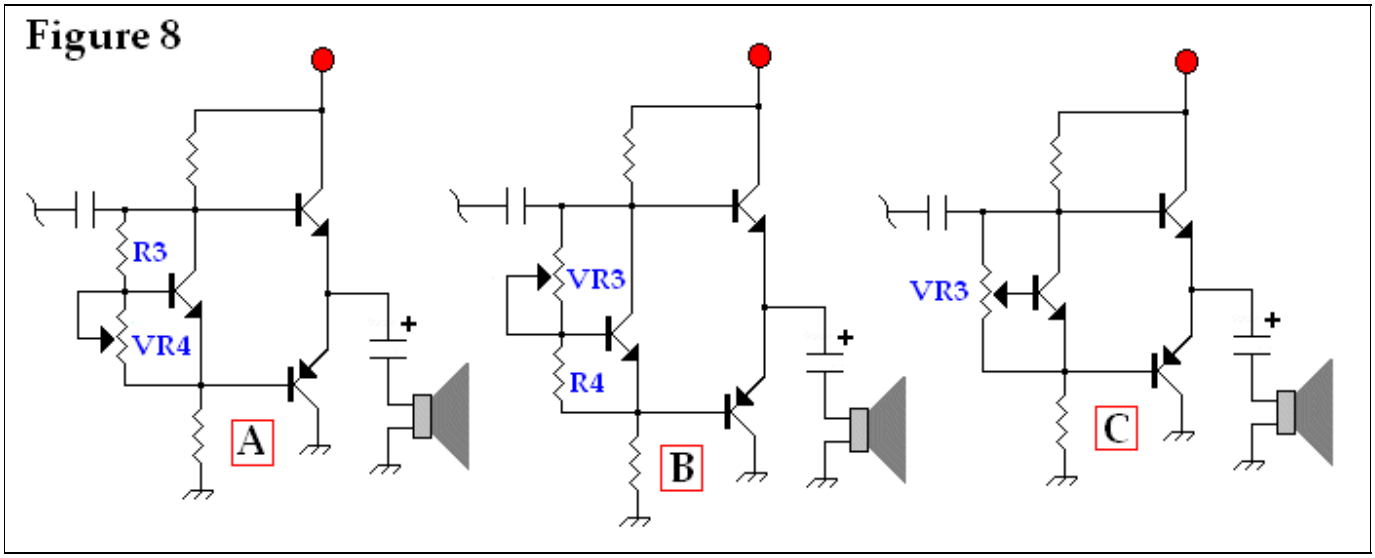


Figure 7 shows the best way to bias our complimentary pair. Our familiar 10K-10K voltage divider is kept, but a transistor Q3 with its own biasing resistors R3 and R4 are added. You might think of R3 and R4 as a voltage divider within a voltage divider. Q3 is referred to as an amplified diode or DC level shifter. It often receives local thermal feedback from the power follower output transistors. This usually involves mounting Q3 on the same heat sink as the finals. If the output transistors heat up, so does Q3 and this results in a a smaller voltage drop across Q3 which translates into less forward bias to Q1 and Q2. Within limits, Q3 with its own base-emitter junction provides variable forward bias for the output transistors.

Shown above is the breadboard of the Figure 6 circuit built on scrap of copper clad board. Transistors were 2N3904 and 2N3906 types, diodes were 1N4148. The capacitor and resistor to the right were a low pass filter (10 ohm and 0.1 uF) to stabilize the output. The unseen speaker was connected to the red and green wires.

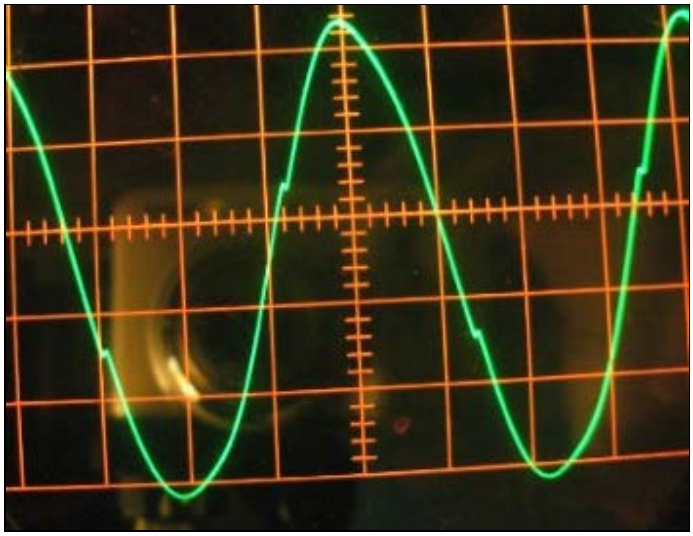


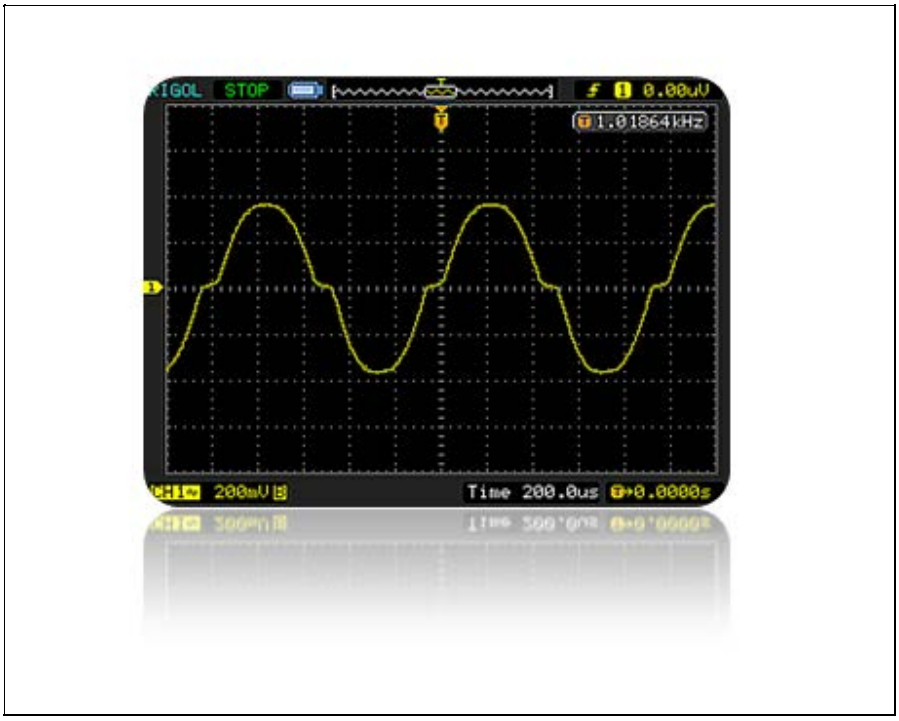
In practice, either R3 or R4 is often replaced with a trimmer potentiometer or a trimmer potentiometer is used instead of R3 and R4 and sometimes R3 and R4 are not of equal value. Shown above in Figure 8 are 3 variable bias topologies for Q3 that I have used. In some cases you will notice that the builder places a fixed value resistor or even a diode in series with the potentiometer in circuits like A or B. Using a potentiometer allows precise adjustment of the quiescent bias current and the ability to dial in the lowest crossover distortion possible. You can set the bias current using any combination of an oscilloscope and signal generator, a voltmeter, an ammeter or possibly try do it by ear when listening for and removing crossover distortion at low volume levels. The procedure I have read to adjust the bias by listening is as follows: Allow some low level signal through the amp so you can just hear it in the speaker. Turn the potentiometer from 1 extreme to another until crossover distortion is heard. Move the pot in the opposite direction until the crossover distortion disappears.
From my limited experience; in some amplifiers under 2 watts or so, you may not hear much of an audible change in crossover distortion when adjusting the bias control potentiometer, so the listening method is not useful in certain cases. It is worth mentioning, that crossover distortion sounds awful and you can usually hear it in amplifiers that are under biased.

Many builders just have a multimeter. In this case, measure the voltage drop across Q3 (the amplified diode) and ensure that is a least 1.1 volts and then slowly adjust the bias up or down from that point. Ultimately, you may have to just make the final bias setting by deciding what voltage drop across Q3 and/or what complementary pair quiescent current you want to establish. It is really not that difficult. *Whatever method you use,*

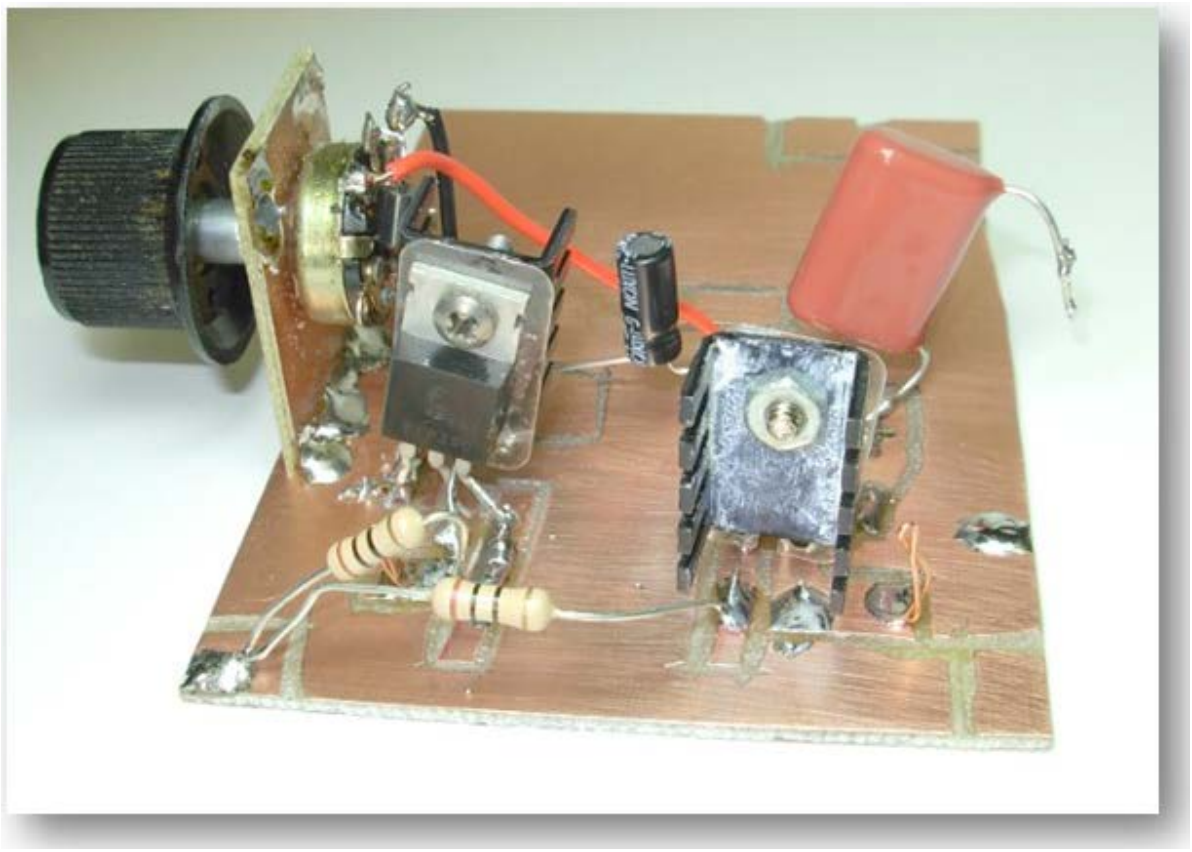
always re-check the Q3 voltage drop and amplifier bias current with no input signal to inform yourself of what is happening.
I am uncertain of the best method to measure the amplifier quiescent current, however I normally measure it using an ammeter connected in series with the emitter of the NPN transistor of the complementary pair.

Shown to the right and below are more images of what crossover distortion can look like in an under-biased power follower stage. The *right* output waveform also contains a little harmonic distortion, however that's a separate issue. I listened to this amp when connected to a speaker and music audio source; the audio had a noticeable "grungy" distorted sound. As mentioned, crossover distortion sounds terrible. The bias to the power followers (a complimentary pair of two 2N3906 and two 2N3904 transistors set up as Darlington emitter followers) was increased and the crossover distortion disappeared. A post bias adjustment audio listening test confirmed that the crossover distortion was gone.

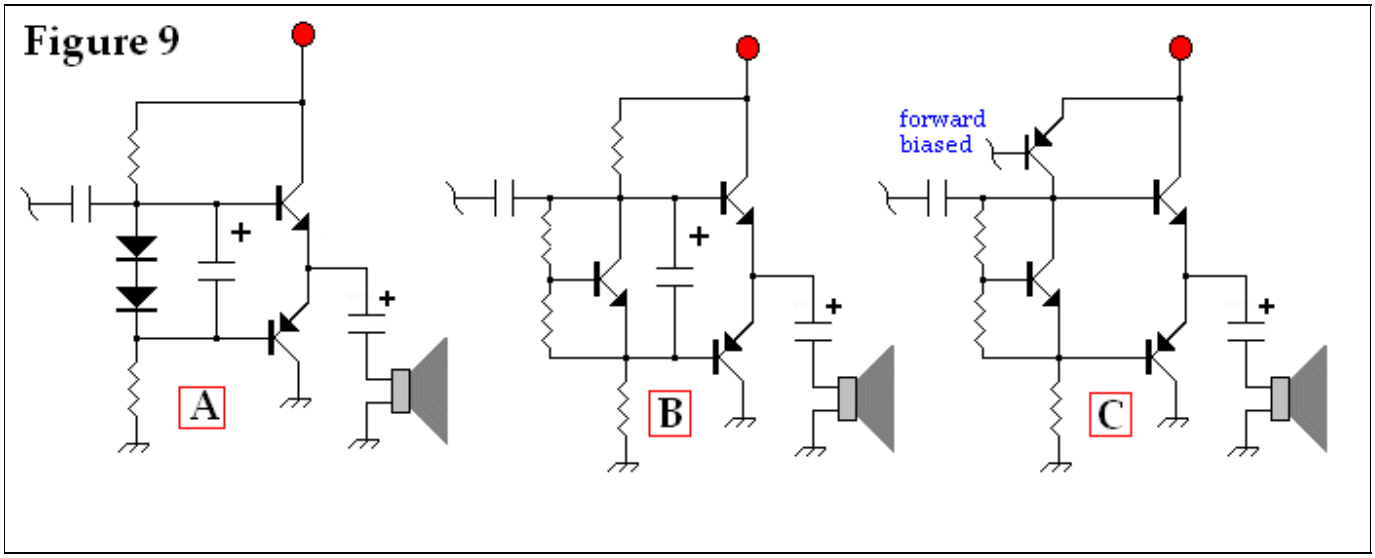




Click for a Russian language mp3 audio file. Vladimir (Volodya), a fellow builder in Ukraine, wanted an A - B comparison of Class AB versus Class B (cross-over distortion). In this audio file, I tweak a potentiometer biasing a pair of power followers to give contrast between the 2 amplifier classes. Under biased AF amps sound terrible in any language! The audio source was a cassette player. The speaker output was recorded, digitized and stuffed into the mp3 file.

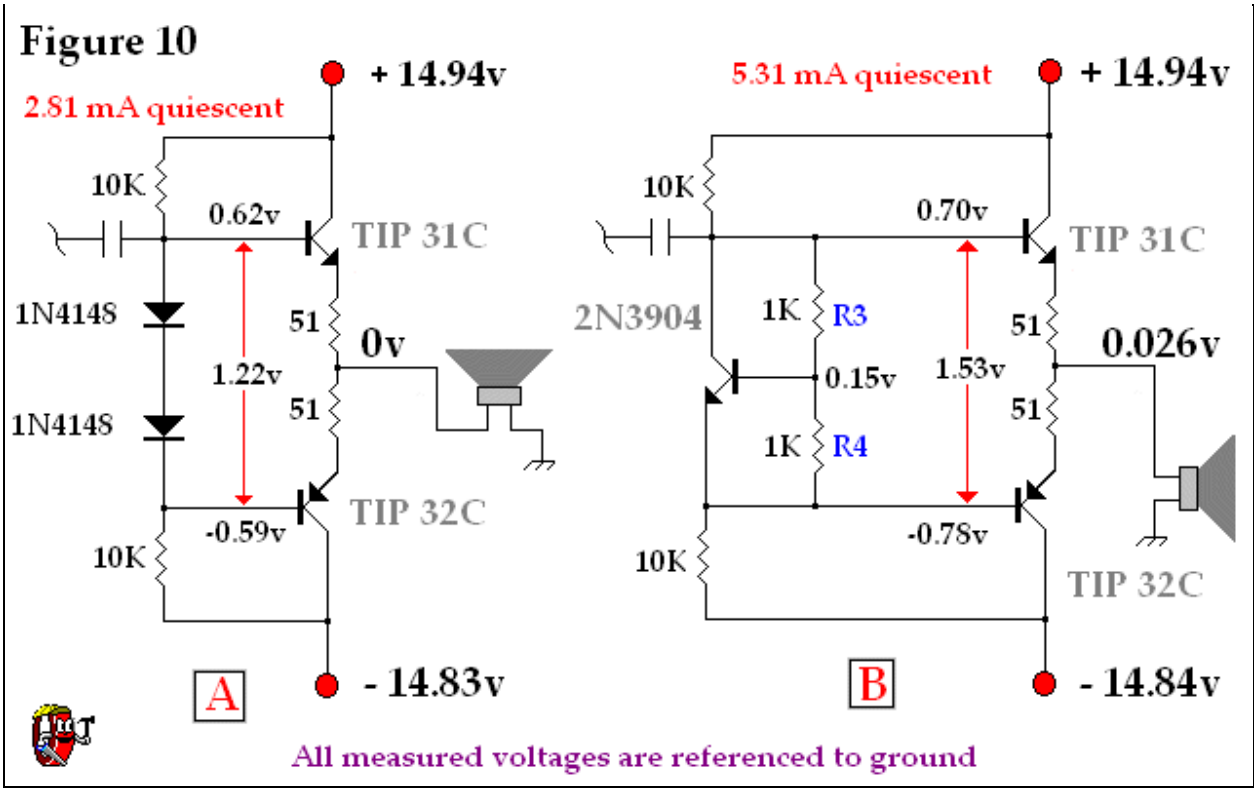


Shown above is a bread board of a complete amplifier utilizing a 10K pot to vary the bias on the amplified diode *(see Figure 8 c)*. With a 12.22 volt power supply, turning the potentiometer from one extreme to the other varied the current draw of the amplifier from about 0 to 95 mA. The average quiescent current draw of a properly biased **single complimentary pair** was somewhere between 5 and 10 mA in my bread boards.

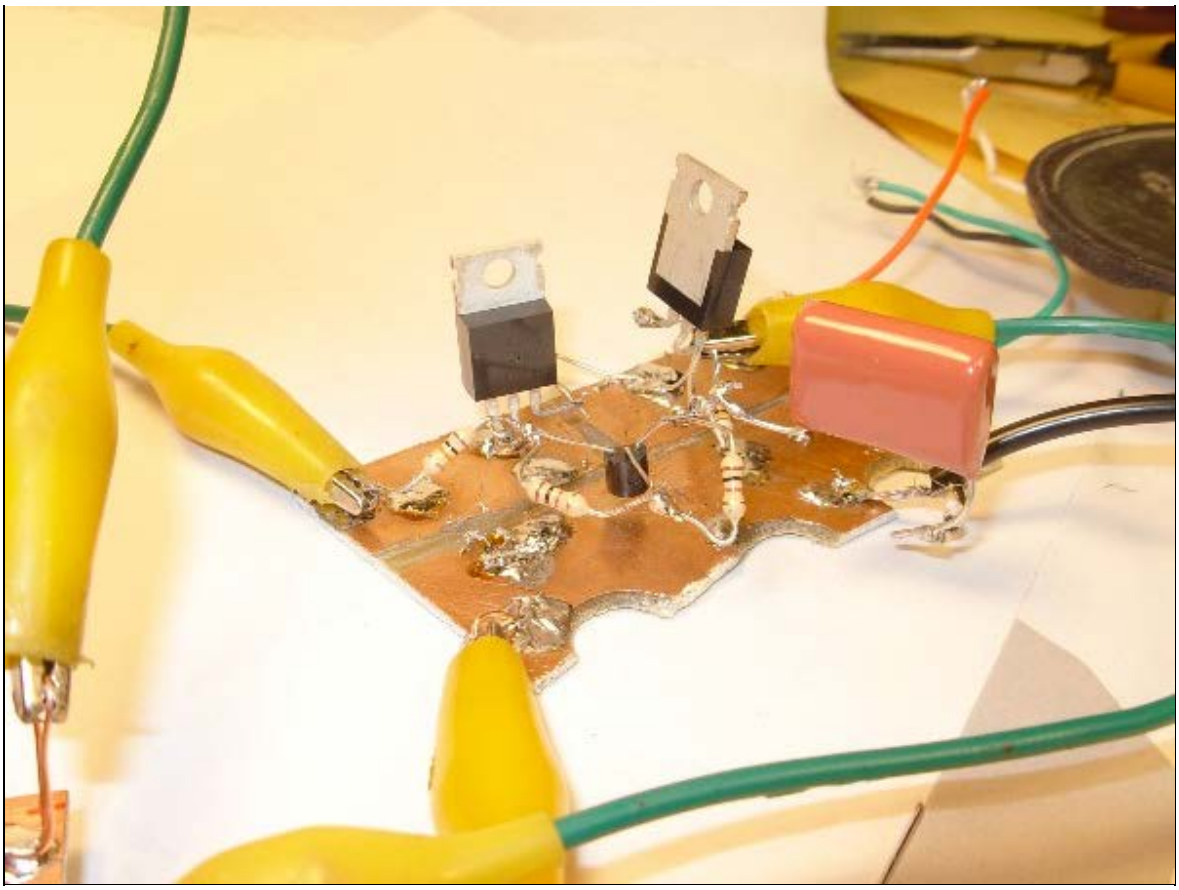


You may see a capacitor inserted between the output transistor bases as shown in Figure 9A and 9B. I have seen capacitor values from 4.7 uF to 100 uF used and the value is not critical, however, from my experiments, I have learned it is mandatory. This capacitor serves to keep the bias voltage constant as the AC signal swings up and down. Some engineers refer to the amplified diode an **NPN shifter bias amplifier** or a **level shifter**. Its function is to charge up the capacitor between the bases of the power follower NPN and PNP pair to a voltage difference that establishes the quiescent current.

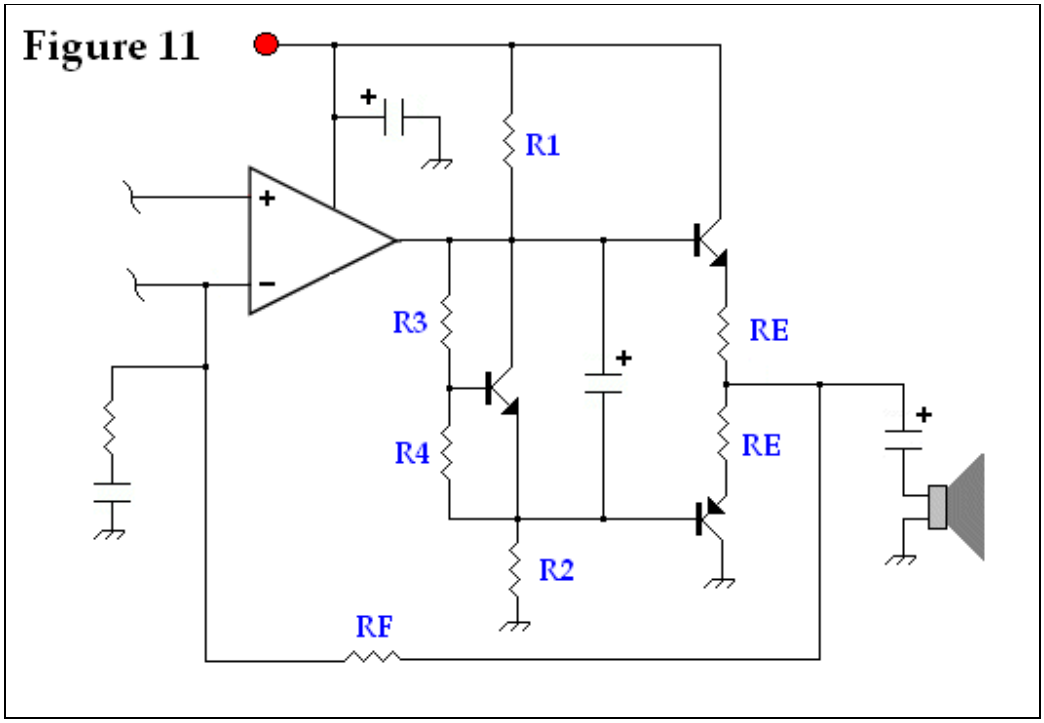
In 9C, R1 has been replaced with a PNP transistor which is usually forward biased by another transistor. You may observe any number of variations of the basic biasing circuit presented in Figures 6 and 7, including 3 or more small signal diodes, 2 amplified diodes, current sources, feedback loops and more. although the techniques vary, the authors are still just biasing the complementary emitter followers to achieve low crossover distortion, stability and/or thermal tracking.



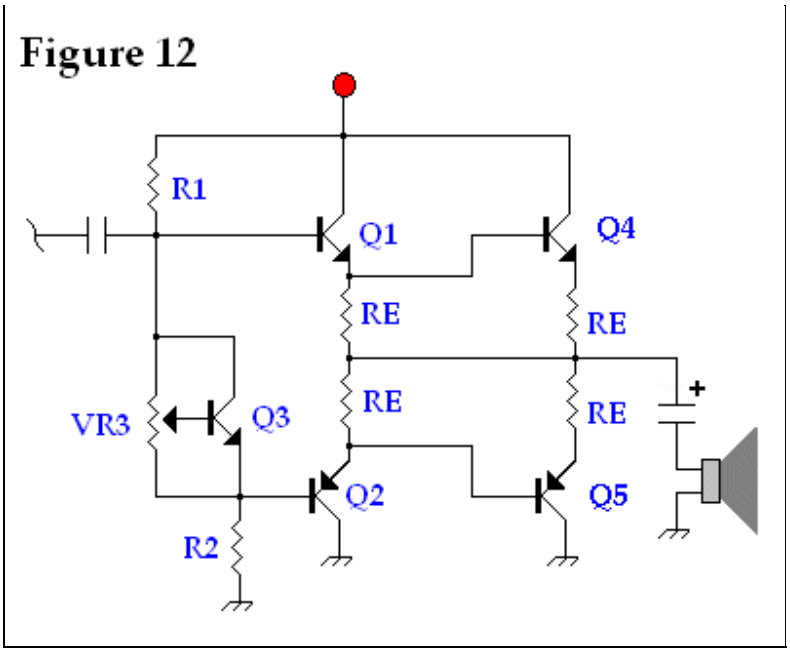
Shown above in Figure 10 are 2 amplifiers using a split power supply. The split power supply offers increased headroom due to a greater AC voltage swing as well as increases the available RMS output power without using super high AC power transformer secondary voltages. In addition, the split supply works well with op amps and if desired, enables you to reduce the number of coupling capacitors by allowing direct coupling of the preamp and speaker to the power amplifier. Coupling capacitors alter frequency response and perhaps may present phase shift issues. In some cases, we as builders use coupling capacitors to provide effects such as high pass filtering, however in Hi-Fi amps, enhanced low frequency response is usually desired; which necessitates the use of high value coupling caps in single power supply amplifiers. In split supply amps, the choice of using a coupling capacitor or not is available to you. In the Figure 10 a and b circuits above, the speaker is directly coupled to the complementary emitter followers output. Note that the voltage at this point is 0 or nearly 0 volts. For any given power supply voltage you chose (split or not), please ensure the amplifier components can handle the current and subsequent heat when a signal voltage is applied. This topic is out of scope. Build and measure...build and measure...



Shown above is a breadboard of the Figure 10b circuit. Additional experiments using even higher voltages were also performed, hence the moderate power TIP transistors were utilized. I burnt up four 2N3904/6 transistors performing many experiments with biasing over 3 nights. Some of the outputs of these experiments will be presented in future projects.



Shown above in Figure 11 is a complimentary emitter follower pair directly coupled to an op amp. The amplified diode and its biasing network is inside the op amp feedback loop. There are examples of this circuit in EMRFD and also on this web site. In single supply powered op amps, it is possible to omit R1. An example of this may be found in Figure 12.30 in EMRFD. Using a low noise op amp such as the NE5532 to drive your power followers can give outstanding results.



Shown above in Figure 12 is another theoretical power amp which illustrates the building block aspects of the simple stage we have been discussing. Q4 and Q5 are cascaded with Q1 and Q2 to build up the current (Darlington emitter followers). Such an amp could have several watts of output power depending on the supply voltage. The emitter resistors on Q4 and Q5 are often 0.47 to 1 ohm power resistors.

References

EMRFD Although Rick Campbell and Bob Larsen contributed chapters and circuits, the principal author is Wes Hayward. It amazes me that any human being could know so much about electronics and is so willing to share his knowledge. Respect.

Henderson, John. Electronic Devices Published in 1991 by Prentice Hall

Oleksy, Jerome E. Practical Solid-state Circuit Design Published 1974 by Howard W. Sams and Co.,Inc

Slone, Randy G. Understanding Electricity and Electronics Published 1996 by TAB Books-McGraw Hill

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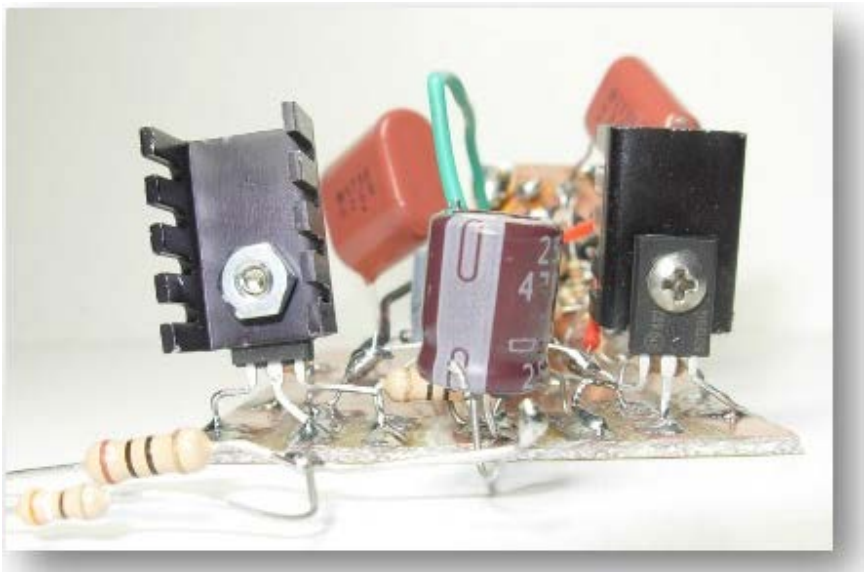
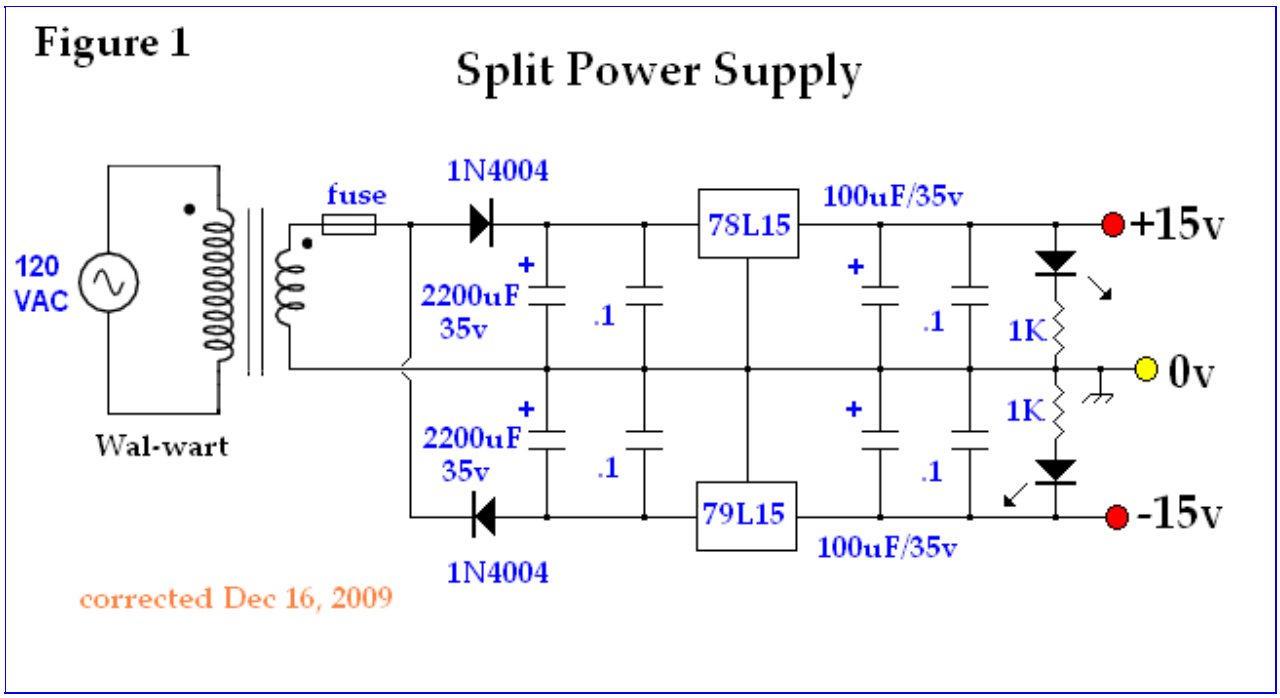
Low Power Audio Amp Experiments

Introduction

This web page contains some experiments on simple, low power, speaker output audio amplifiers. Presented are ideas, some measurements and examples of audio amplifiers which will likely sound better than the IC audio chips commonly seen in many receiver projects. This web page is a follow-on to this one and is a completely new area of experimentation for me. Audio amps were built using both split and positive power supplies. In all cases the complimentary power followers were driven by an op-amp. I tried building some power amps using discrete transistor differential amplifier stages with current sources as the driver, but the noise performance and simplicity of the NE5532 or NE5534 op-amp was superior.

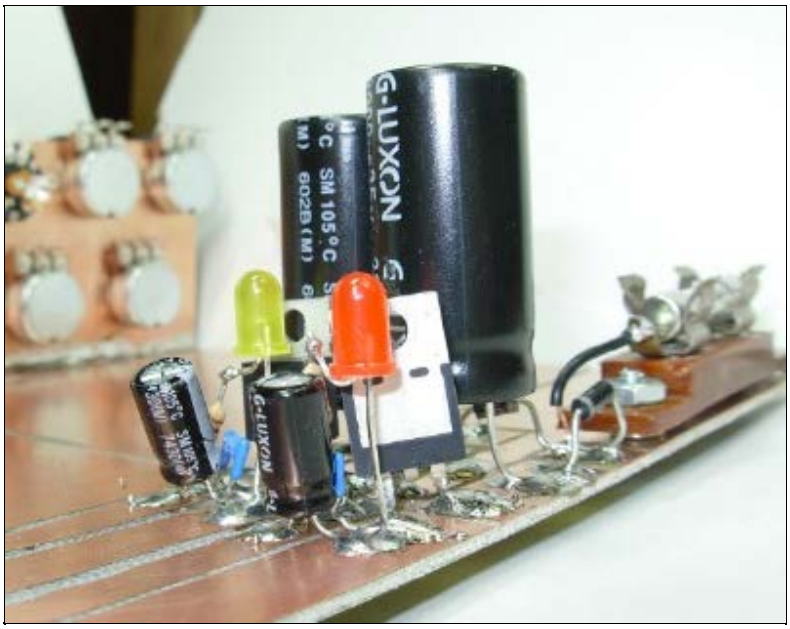


Split or Bi-polar Power Supply Audio Amplifiers



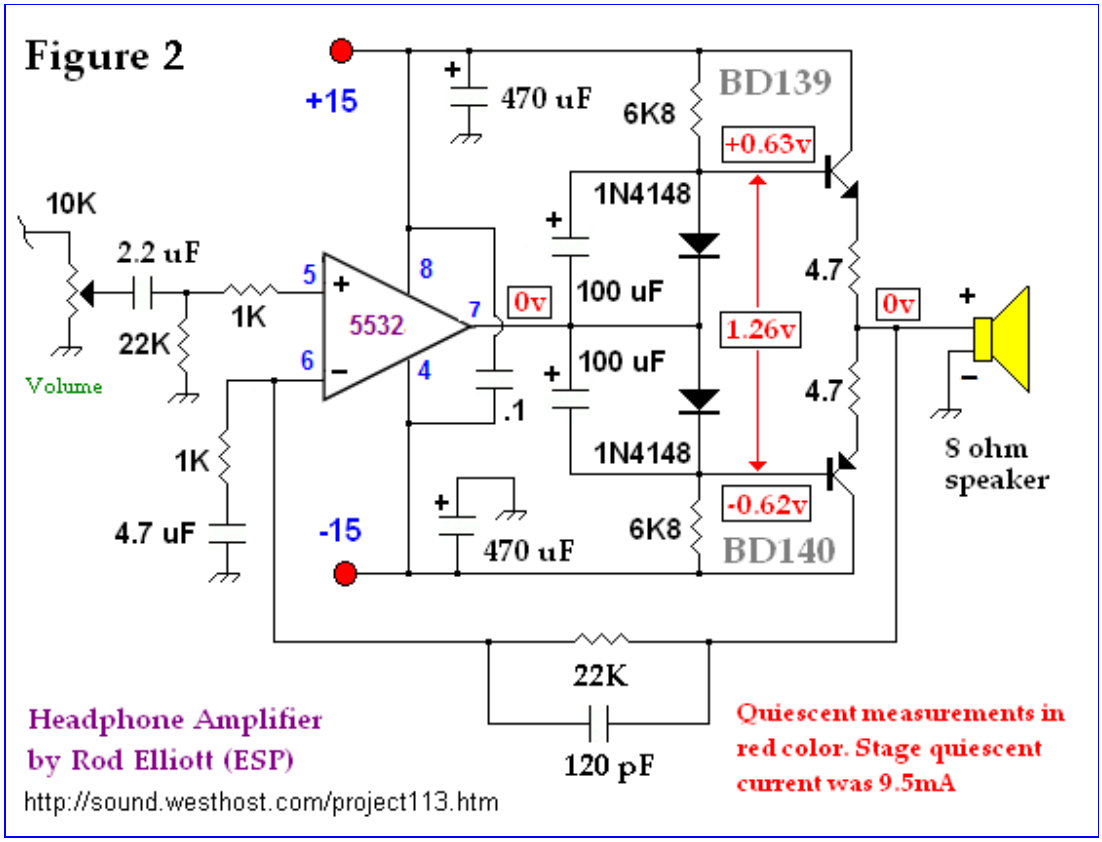
In order to build up some split power supply amplifiers, a basic power supply was constructed and the schematic is shown in Figure 1 above. I

found it was essential to regulate the voltage or hum would appear on the output. Choose a standard value fuse that is rated somewhere just above the maximum current you measure. I used 2 different AC output power transformers which were in the 18-24 volt, 375 mA to 1 amp range. The LEDS are strongly suggested. They inform you when there is power applied and their relative brightness will also often fall when higher current is being drawn on one side or the other. This alerted me to an accidental solder bridge to ground on more than 1 occasion.



The split power supply is shown in the photograph to the left. The retro Bakelite fuse holder is from an old tube audio amplifier. If you are wondering why the copper clad board is so large, the power supply is part of a future project. Some builders would use even greater value filter capacitors than those shown. Heat sinks on the voltage regulators are required for supplying DC to higher power amplifiers.

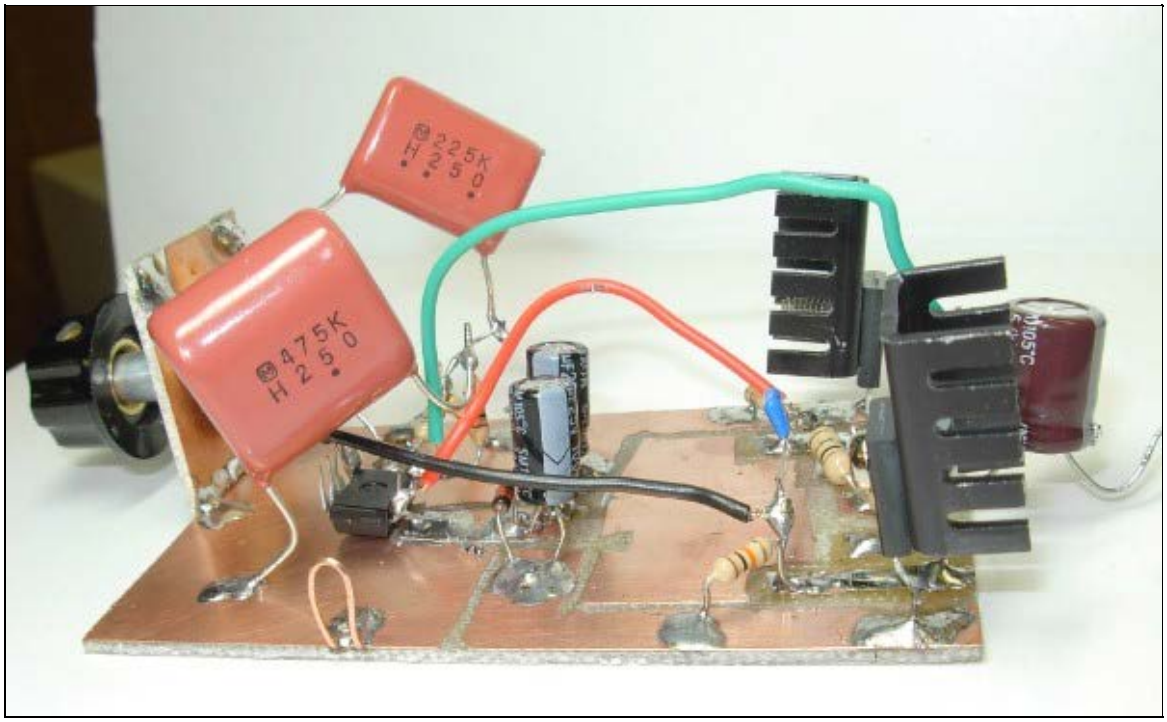
Rod Elliot Headphone Amp



The first amp I built is shown in the Figure 2 schematic above. This amplifier was designed by Rod Elliot and is used with his permission. Rod's ESP web site is a virtual treasury of audio design information. If you are into understanding audio design, visiting his web site is strongly recommended. Rod sells printed circuit boards for all of his circuits if you prefer this building method . Note I have made some minor modifications to some part values. although primarily designed as a low distortion, high power, headphone amplifier, it drives an 8 ohm speaker very well. I was able to drive this amplifier as high as 0.68 watts average power with a pure sine wave output during analysis. Power measurement is discussed in the next section. Note that on this web page, I quote the entire stage quiescent current. Since the op-amp and the

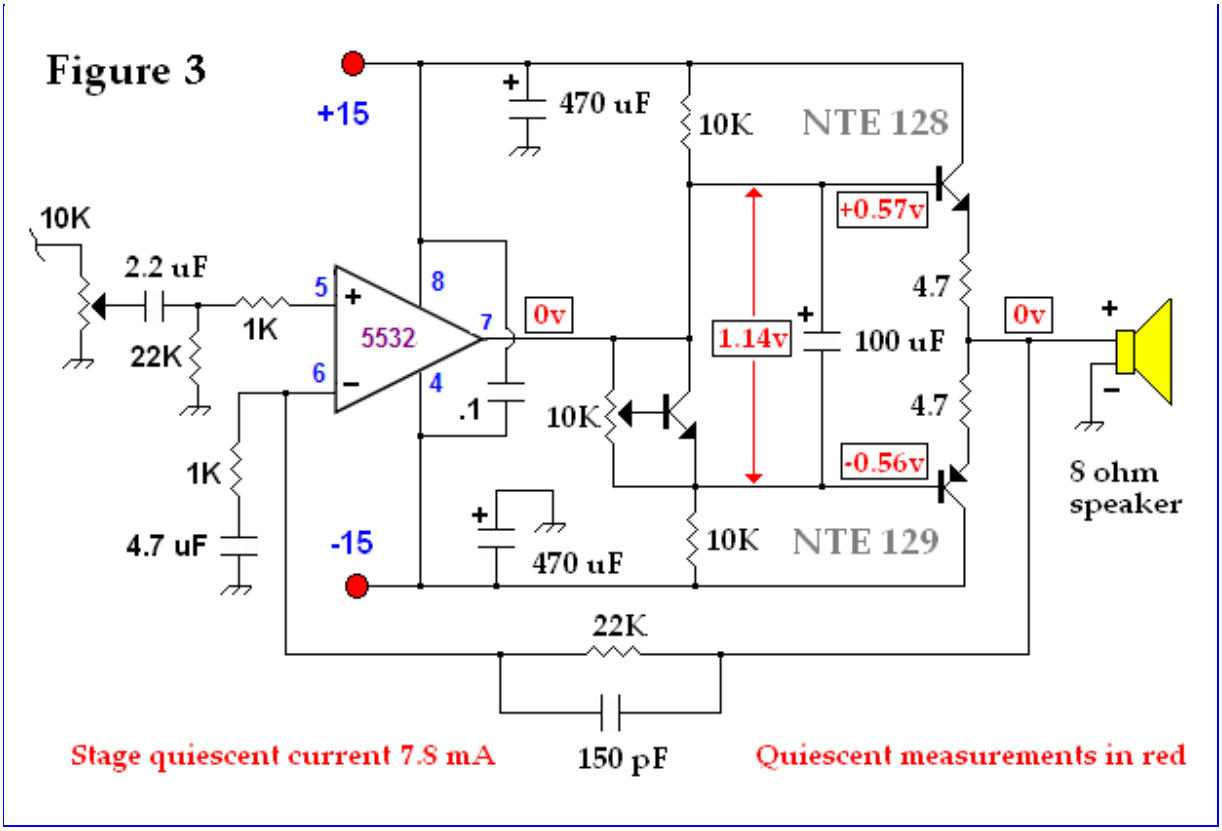
2 (or more) power followers are a "package", it is a lot easier to just measure the current at the power supply lead(s) of the stage than unsolder and lift up a transistor lead. In this case, with no input single, the stage current was about 9.5 mA. I did check and about half the current is going to the op-amp with the other half to the transistor pair. This is a wonderful sounding amplifier and Rod has an entire web page devoted to it, so I will not comment further.

There are a variety of suitable transistor pairs for audio power amplifiers depending on the power output you are choosing. I stock just a few; BD139-140, TIP 41C-42C, NTE 128-129. The higher beta 2N3904-3906 or 2N4401-4403 pairs worked well in the low power, single power supply amplifiers shown on this web page. I also performed some higher output power experiments which required the TIP and BD transistors and these are not shown.



Above photograph. A breadboard of the Rod Elliot headphone amplifier. This early version had a temporary output capacitor. When first testing a new circuit that has a direct speaker output, it might be a good idea to temporarily use an output capacitor until you measure your voltages and current and feel your transistor temperatures. This will save your speaker if you made a big mistake and/or blow up the transistors when you first power it up.

Amplified Diode Biased Audio Amp



Above in Figure 3 is a split power supply audio amp using an "amplified diode" to control the bias. The bias transistor was wedged between one of the output transistors and a piece of copper clad board to allow thermal tracking. The 10K bias control resistor was a trimmer type suspended over the copper clad board in most of my bread boards. Usually, you just need to set and forget about this resistor after initial set up. I adjusted the bias by watching in my oscilloscope with a low level, 1 KHz sine wave connected to the input. I measured the various voltages and stage current at quiescent and have indicated these values in red for learning purposes. The bias current range was 7.2 to 154 mA when turning the 10K trimmer pot from 1 extreme to the other. The maximal clean output average power of this amp was 0.78 watts. I used press on heat sinks for the NTE128-129 pair and they ran quite warm to touch. These TO39 type packaged transistors are somewhat difficult to heat sink compared to the TIP/BD transistor packages where you can just bolt on a heat sink of any size that is required. Please remember that the metal tab on the TIP and BD transistors is connected to the collector terminal.

In the above photograph is my first bread board of the Figure 3 amplifier. This particular version had TIP transistors, a 4 ohm speaker and an 470 uF output capacitor. Note the full size 10K bias control potentiometer on the left hand side. This was purely a experimenter's bread board, but it sounded amazing when listening to music through it.

Harmonic Distortion and Measuring Output Power

Average Power Formula

Applies to a continuous sinusoidal tone into a purely resistive load

$$P = \frac{\text{Peak Voltage}^2}{2 \times \text{Load Resistance}}$$

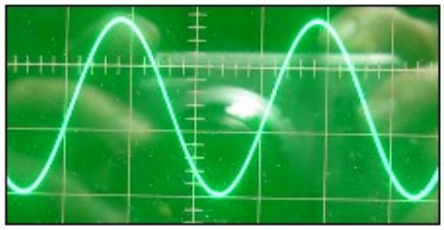


Figure 4

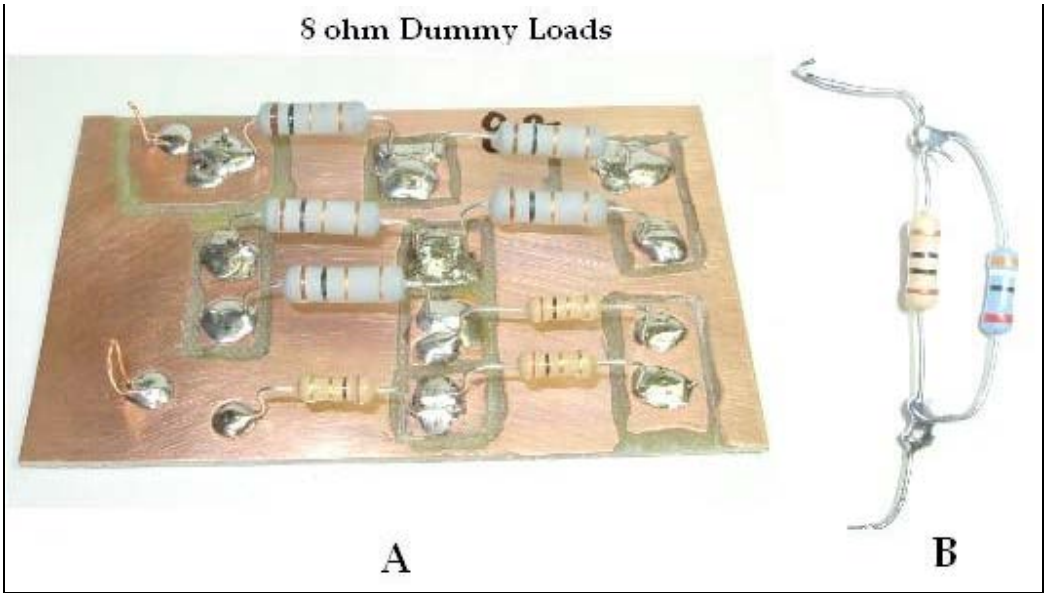
In Figure 4 is the formula used to calculate the average power of the circuits on this web page. For example if you measure 6 volts peak to peak on the oscilloscope, (3 volts peak voltage) and your resistive load is 8 ohms, the average power is 560 milliwatts. At any point in an AC waveform there is power and it may be reported using a variety of ways. Was it clean? distorted? a peak value? an RMS value? - often it is unclear.

To be clear, I measured the peak voltage on a pure, undistorted sine wave into an 8 ohm resistor. Stated power values are the mean sine wave power calculated with the formula shown. See this somewhat controversial link for details. You may not agree with my methodology, however, it allows you to compare the circuits on this web page. If you really must know the peak power, multiply the stated average sine wave power by 2. I will leave the power measurement and calculation debate up to scholars; as a lay-person, I need something simple.

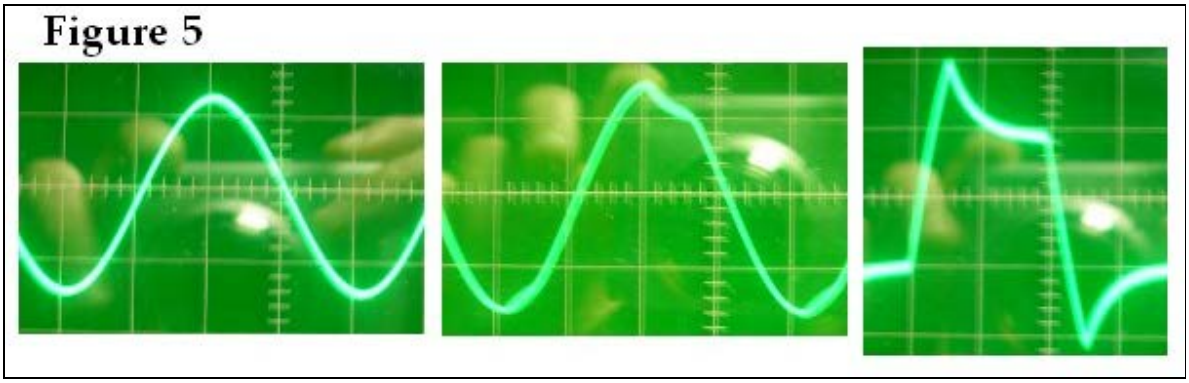
The bench voltage measurement was as follows: The amplifier was connected to a 1 KHz pure sine wave generator and the 10K volume control pot was advanced just until any sign of distortion of the amplifier output sine wave appeared. Voltage measurement was taken at the point just before distortion occurred.

It is difficult to photograph a sine wave without a tripod. Motion, the angle, light reflection and jpeg graphic compression all wreck the perfect sine wave. In the Figure 4 graphic above is a typical 1 KHz output waveform from my power amplifiers (squeaky clean) at the amplifiers maximum average power level.

To the right is a photograph of my AF signal generator. This is an old, tube device but the output sine wave is beautiful. I did not perform spectrum analysis with a computer audio sound card program and will leave this up to audiophiles. These audio amps sound great; especially when compared to the IC audio power amps that many of us tend to use in our receivers.



In the above photograph is **A**, an 8 x 1 ohm resistor load and **B**, an 8 ohm load made from parallel 1/2 watt 10 and 39 ohm resistors. In dummy load **A**, I used 5 two watt metal film resistors plus 3 half watt resistors. In the future, I will obtain 3 more 2 watt resistors and replace the 1/2 watt resistors for a 16 watt rating. For a quick resistive load, **B** is the way to go for most of the circuits on this web page.. You can make a 4 ohm load from parallel 4.7 and 27 ohm resistors. In truth, a single resistor or any combination of resistors adding up to the desired load R value will work.



In Figure 5 above are some scope waveforms ranging from mildly distorted to full-on dirty.

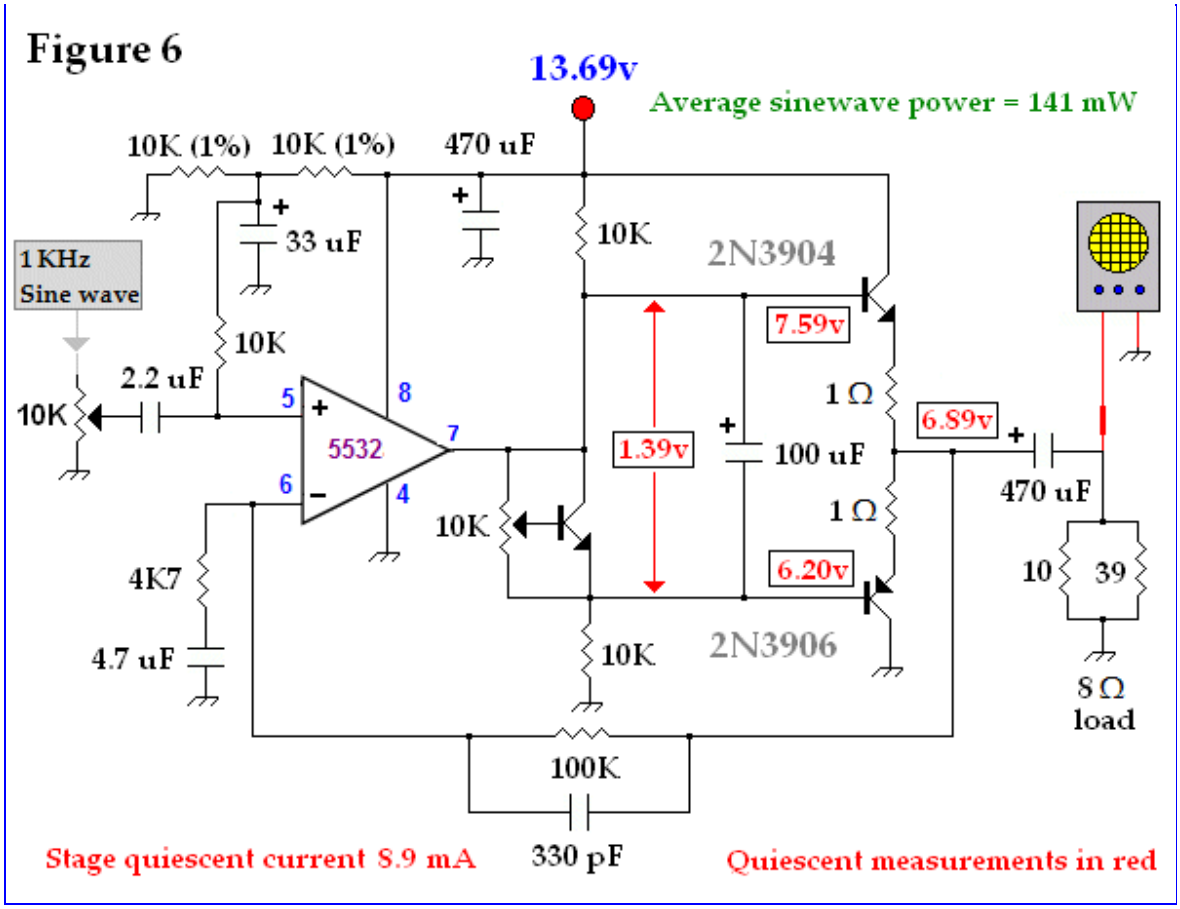
Power Amplifier Concerns

Although the amplifiers on this page are 0.15-0.8 watts or so, they can consume relatively large current compared to the usual voltage amplifier circuits we build. Some potentially helpful tips to help keep away ground loops, oscillations and thermal run away are suggested as follows:

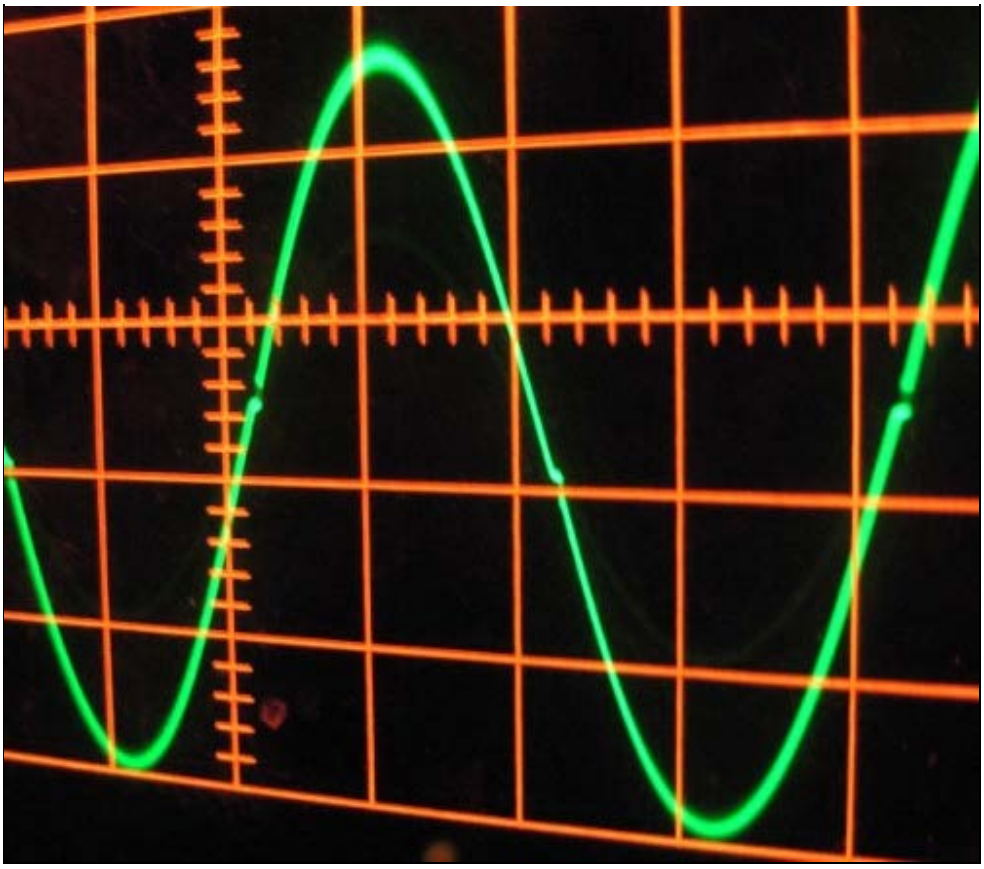
✓ Connect your negative speaker terminal directly to the AF power amp (do not use a common ground for the negative speaker terminal).
✓ Use big power supply line bypass capacitors (no 10 uF caps here)
✓ Keep your audio amplifier copper clad board separate from the rest of your circuit boards and star ground it to your main power supply ground point.
✓ Use heat sinks on your final transistors and voltage regulator(s) when you go for bigger power
✓ Watch your layout - keep the output away from the input etc.
✓ Watch your emitter resistor power ratings in "higher wattage" amplifiers. Burning resistors stink.

Single Power Supply Audio Power Amps

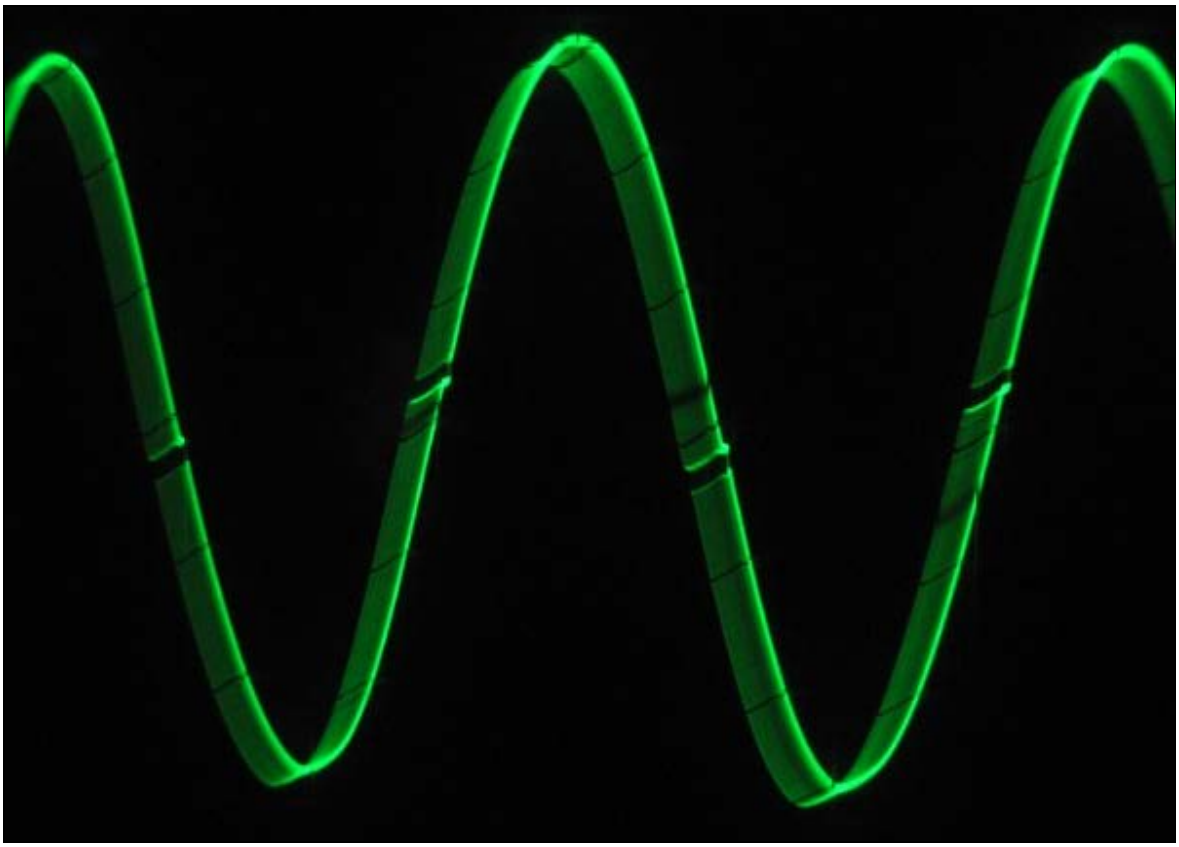
Since most 12 volt power supplies are actually closer to 14 volts; these experiments were performed with a typical radio bench DC power supply at 13.69 volts.. Figure 6, 7a and 7b represent evolving experiments aimed at obtaining greater output power.



Shown above in Figure 6 is the fundamental design using one op-amp and 2 power followers. It is shown in an AC output power measurement configuration. The bias current range was about 5 to 100 mA when turning the 10K trimmer bias control pot from one extreme to the other. Maximal sine wave average output power was only 141 mW. Nevertheless, it might be loud enough for some receiver applications.
I connected this amplifier to a VCC of 15 volts. The maximal sine wave average output power was then 220 mW. In all of the single supply audio amps presented , increasing the VCC increased the maximal power output. Driving these amplifiers beyond a pure sine wave output power resulted in predictable harmonic distortion plus the re-emergence of crossover distortion in the output. This was an incredible learning; how could there be crossover distortion re-emerging in a amp that was properly biased to begin with? Increasing the bias current to the maximum level did not remove this crossover distortion. After emailing this question to Rick, KK7B and Wes, W7ZOI, and reading their replies, my best guess was that at some power level, the 5532 op-amp can not provide enough current to properly drive the complementary symmetry pair. The AC current in the output transistors may be limited by the base drive of the op-amp and they were no longer forward biased at the crossover point.

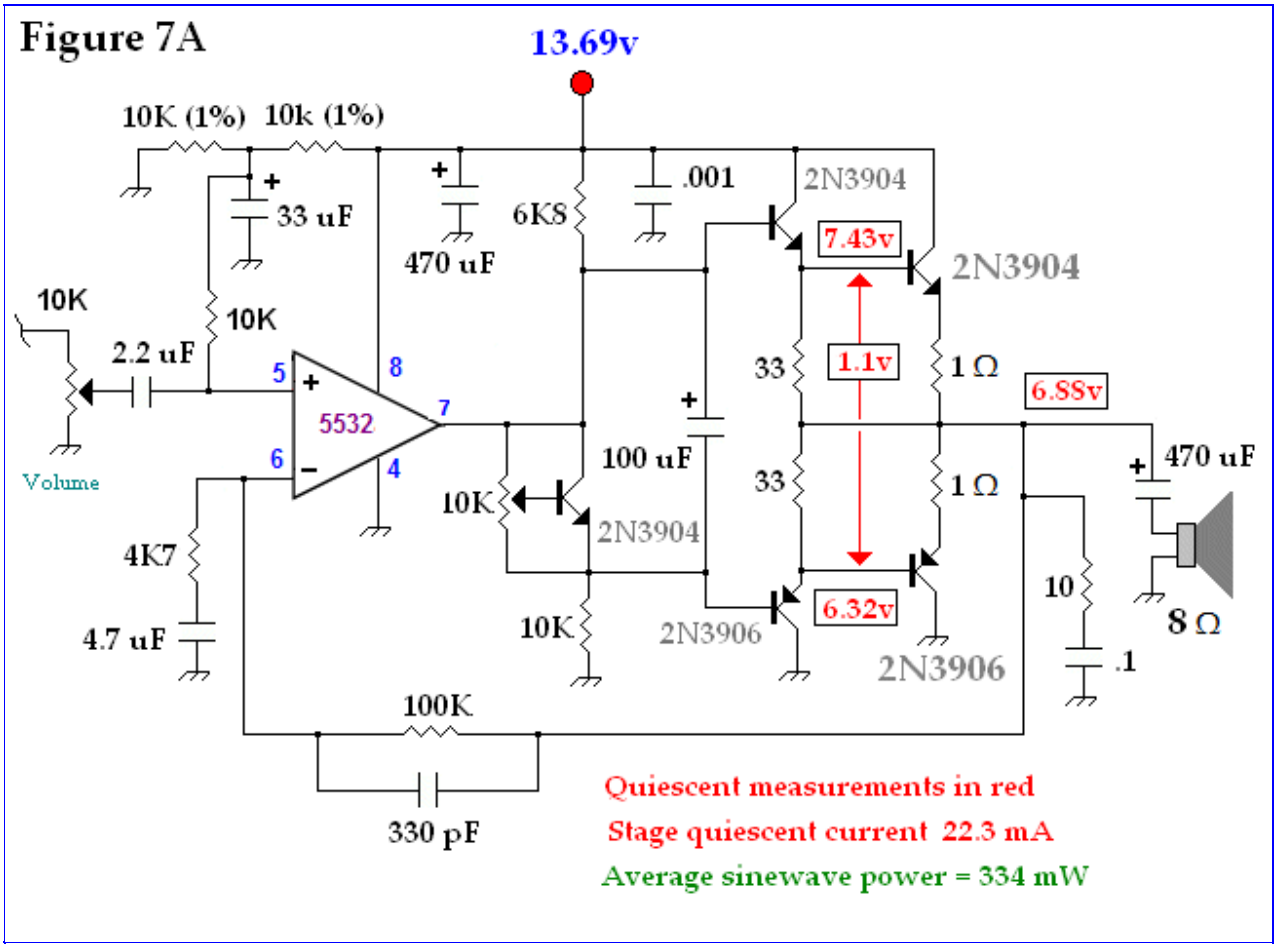


Above photograph. This is the Figure 6 amp driven past the point where the sine wave is pure. Note the crossover distortion blips on the sine wave. The base drive current for the power follower pair all comes from the op-amp. At this point there is likely not enough base drive to keep the base emitter junctions forward biased.



In the above photograph I blacked out the room and photographed the same scope waveform as above while shaking the camera from side to side. This adds some horizontal spreading of the signal and provided more information about what was happening as compared to a single,

clear oscilloscope trace.
I should mention that this crossover distortion blip occurred in all of the Figure 2 to 6 amps when they were driven past the point where a pure sine wave was seen. It is clear that maximal available power from a simple audio amplifier like this **(one NE5532 op-amp plus 2 power followers)** is constrained and thus its application is limited. Greater output power is possible using a split supply per Figures 2 and 3, however, a typical radio project has a 12 volt, single power supply. These basic amplifiers with a single power supply, may be very appropriate for projects such as a compact radio receivers or a code practice oscillator project, but not for applications where you require louder audio.

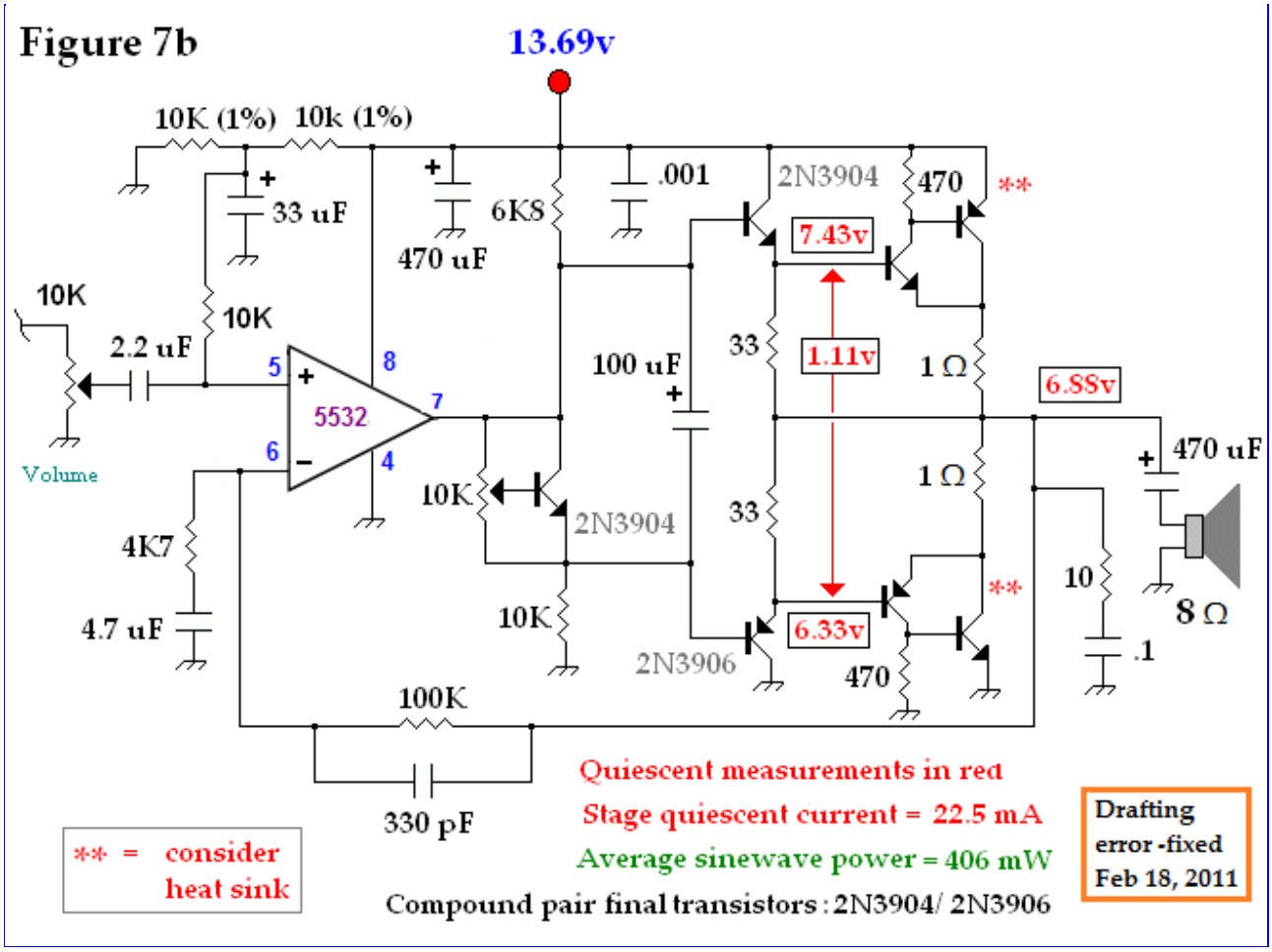


Shown above in Figure 7a is an easy method to get more output power from the Figure 6 amplifier; add another set of complimentary pair current amplifiers. I found 33-39 ohms to be a good emitter resistor value during my experiments. Many hi-fi amp builders will use greater emitter resistor values, however, a design goal was to get more output power from our 12 volt supply. Series emitter resistors are used to improve linearity and operating-point stability. I kept the final power follower pair emitter resistor values at 1 ohm to get maximal output power. An output 10 ohm + 0.1 uF low pass filter was used to help prevent oscillations in view of the low emitter resistor values on the finals.

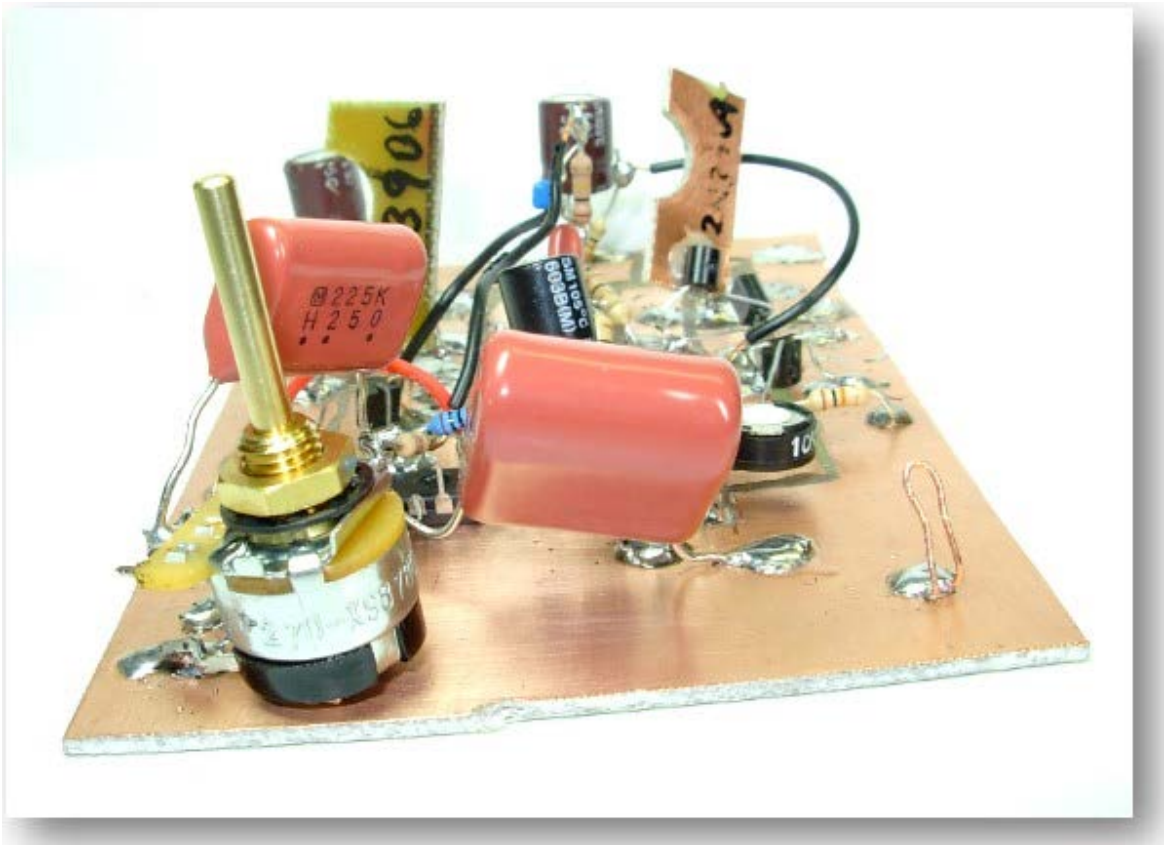
Biasing

The top 10K bias resistor was lowered to 6K8 to facilitate "more linear" setting of the output transistor bias with the 10K trimmer potentiometer. It did not help much. Setting the bias is very delicate procedure and you must turn the screw driver very slowly. In my bread board, the optimal stage bias current was 22.3 mA but anything around 20 mA should be fine. If you do not have an oscilloscope, after ensuring that there is no input signal, connect an ammeter in series with the positive power supply lead. Turn the bias potentiometer with with the screwdriver until you get close to 22 mA. If you only have a voltmeter, the rule of thumb of 1.1 seems to work... Measure the voltage across the final 2N3904-2N3906 bases and ensure the difference is at least 1.1 volts while adjusting the 10K trimmer pot. Personally I do my biasing with an oscilloscope at at least 2 different frequencies on the signal generator, however the for mentioned methods will work okay. This is a popcorn stage and a popcorn web site after all!

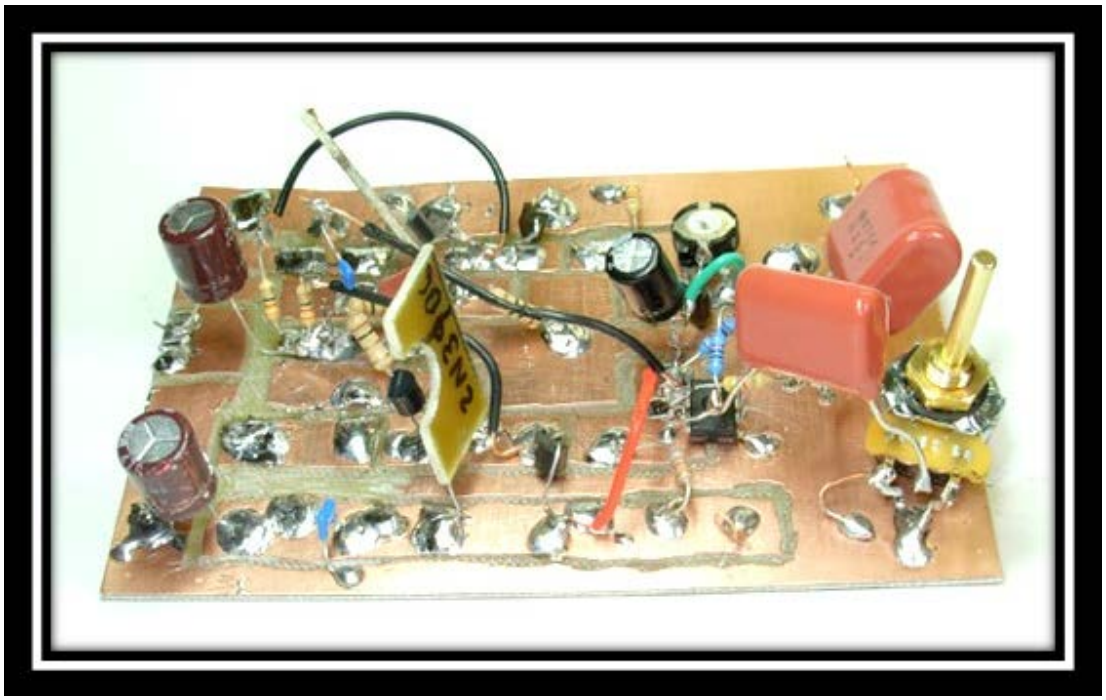
For the lowest potential noise, consider using metal film type resistors in your audio amps and "polysomething" capacitors wherever AC signal is coupled to another component or ground, excluding the output capacitor.



Further experiments to increase output power were frustrating. Finally a compound or Sziklai pair was trialed and increased the average power to over 400 mW as shown in Figure 7b . I used a small piece of copper clad board on the finals for a heat sink, although they really didn't get that warm. Ideally, the amplified diode should also be glued onto one of the heat sink boards for thermal tracking. This amplifier is now in a chassis as a bench reference audio amp for receiver testing.



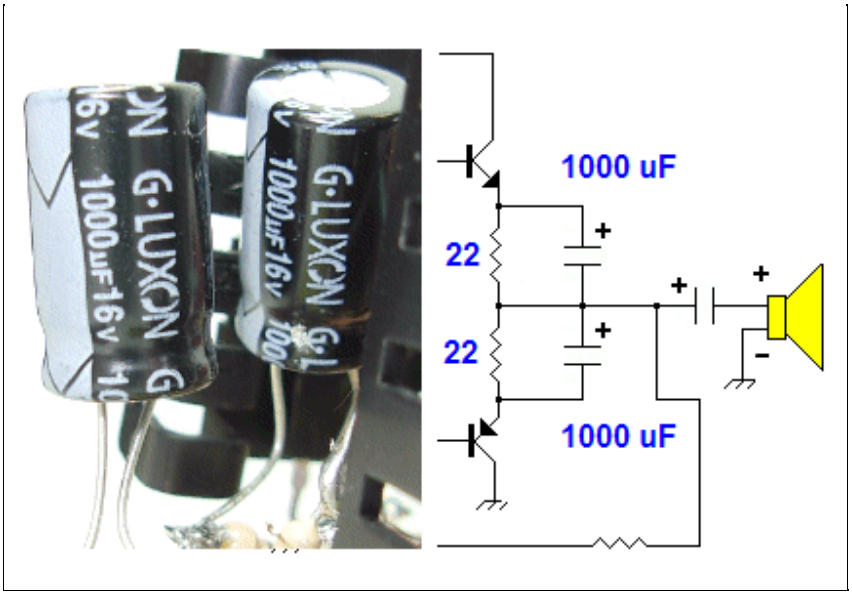
In the above photograph is the Figure 7b prototype. I am using a new miniature potentiometer for my experiments that I bought from Digi-Key. The base has 2 leads which can be soldered right on the copper clad board for easy anchoring and removal after testing.



Another view of the bread board on which the Figure 6, 7a and 7b experiments were conducted.

KK7B Headroom Boosting Emitter Capacitors

We first learned about using large value emitter caps in audio amp complimentary pairs from EMRFD. Experimentation revealed that these capacitors do 2 things: **1.** Can increase the amplifier sine wave headroom and **2.**



Add some low pass filtering. I learned from Rick, KK7B, that he designed his EMRFD amp to achieve low output power, low distortion and lower DC current drain. He desired a clean output audio amplifier for his R2 series of receivers without needing a lot of quiescent current or heat sinks on the 2N3904-2N3906 pair. The caps were added to make the amplifier think it had much lower emitter resistors at AC than the 22 ohm resistors he used in the EMRFD projects. When Rick made measurements and simulations of the amplifier, it was very stable, had low distortion and provided a very nice clean sound at all signal levels, from very weak signals in a few milliwatts of noise, to music driving the speaker. The result is outstanding and Rick's design was the catalyst for my own interest in audio amplifier experimentation.
I performed experiments with these capacitors and found that they increased my amplifier power and head room in some cases, and that the boost is indirectly proportional to the emitter resistor value. With 1 ohm output transistor emitter resistors the boost is generally not that significant. With 4.7 ohm or greater emitter resistors, they can make a big difference and you might consider trialing them for more power and headroom as appropriate. They can also add a nice, warm sound to your audio amp. Refer to EMRFD for numerous examples of this technique.

Popcorn Audio Amplifier

What follows is a popcorn or "poor man's" audio power amp using the 2N3904-2N3906 pair. To meet true popcorn criteria, all of the capacitors used in my breadboard were electrolytic and you can substitute different values from your own junk box. It would be better to use "polysomething" capacitors for the NE5532 pin 5 and 6 signal capacitors if you have them. I normally use a 1 uF to 4.7 uF poly-type capacitor in series with the 4K7 resistor on pin 6. The 270 pF feedback capacitor could be omitted or substituted with a higher or lower value to suit whatever high frequency roll off you desire.
The transistor glue-on heat sinks seen in the 1 bread board photograph are completely unnecessary. This BJT pair were used in other higher power experiments as well. The 22 uF capacitor between the transistor bases is essential from my experimentation. Without this capacitor, the amplifier headroom decreases and crossover distortion occurs. You can use the other half of the 5532 for a preamplifier or use a NE5534 instead. A 741 op-amp would be a horrible substitution. The NE5532 performance is breathtaking considering its low cost. 181 mW is surprisingly loud. All resistors are quarter watt rated. What a fun little amplifier! The schematic is Figure 8.

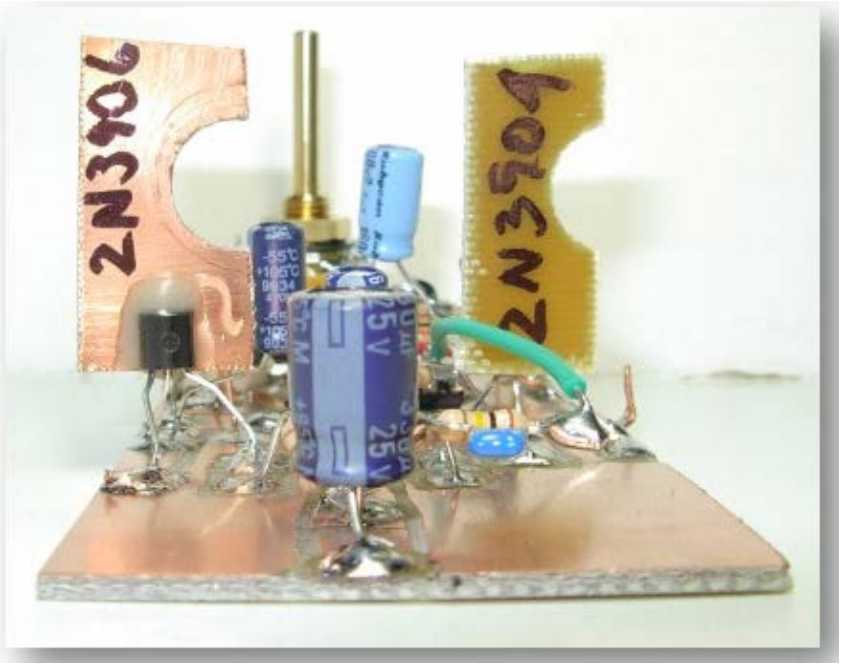
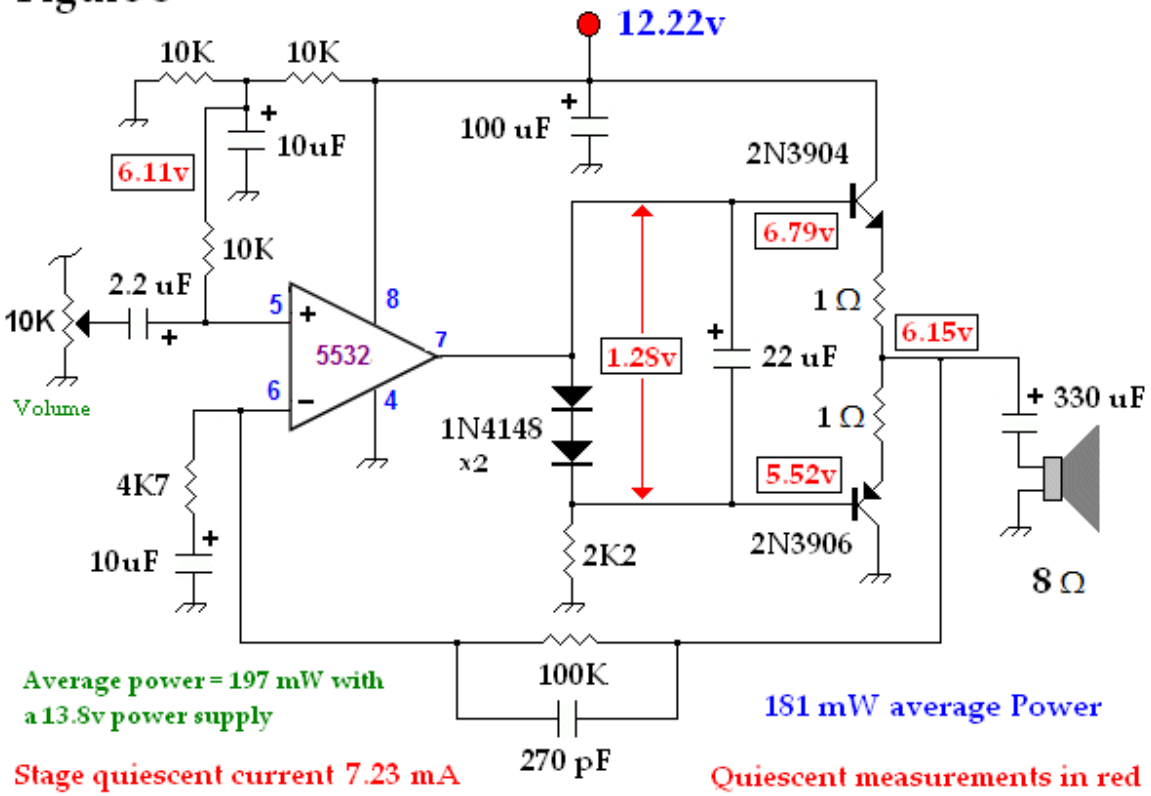


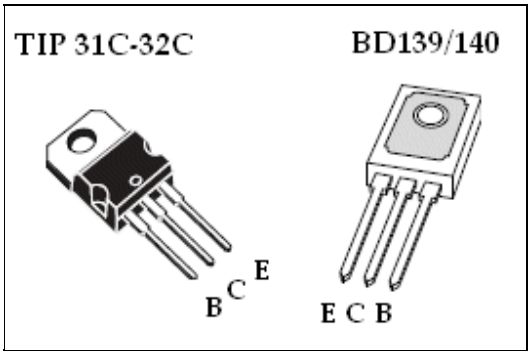
Figure 8

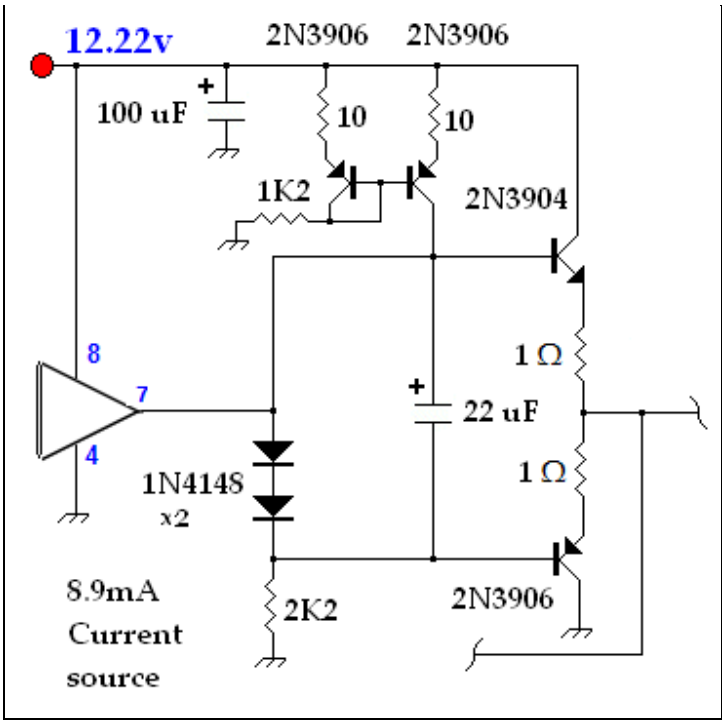

12.22v
10K
10K
10uF
6.11v
100 uF
2N3904
6.79v
10K
2.2 uF
10K
5532
1.28v
22 uF
1 Ω
6.15v
Volume
1N4148
x2
1 Ω
+ 330 uF
4K7
5.52v
10uF
2K2
2N3906
8 Ω
100K
Average power = 197 mW with
a 13.8v power supply
181 mW average Power
270 pF
Stage quiescent current 7.23 mA
Quiescent measurements in red

missing
10K pot

The bread board of the Figure 8 popcorn amplifier using transistors without heat sinks.

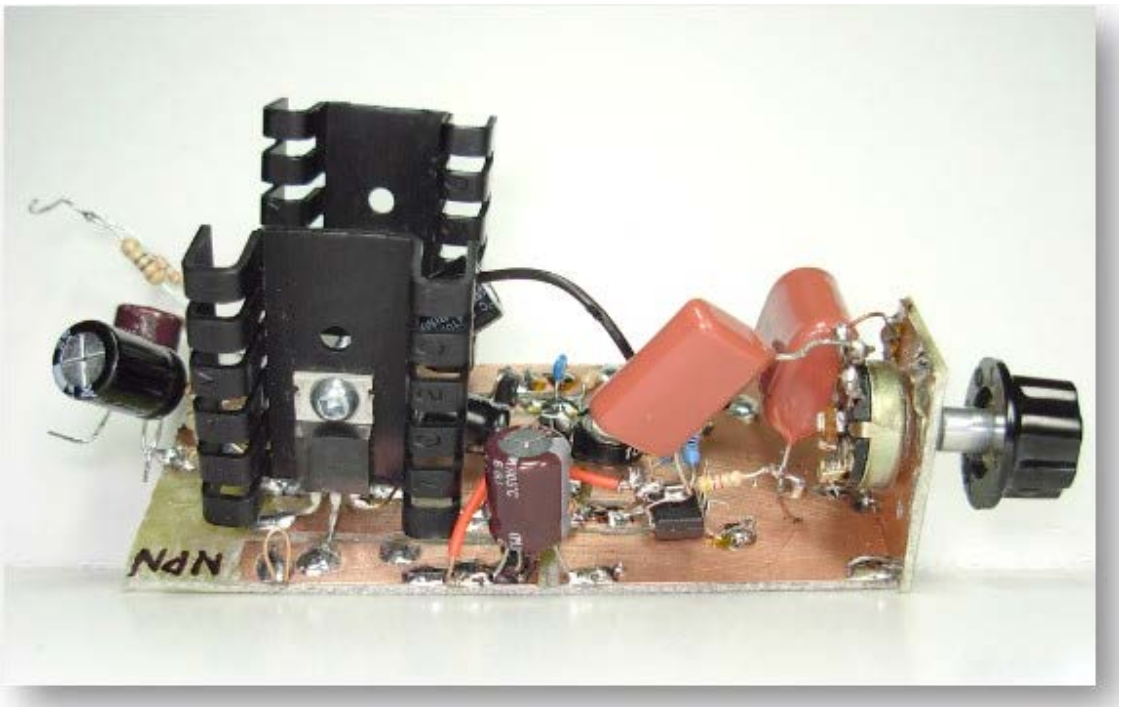
Additional Outputs





I tried putting a current source on the Figure 8 bias and it made no difference to the amplifier characteristics according to my simple oscilloscope, listening and DC analysis.

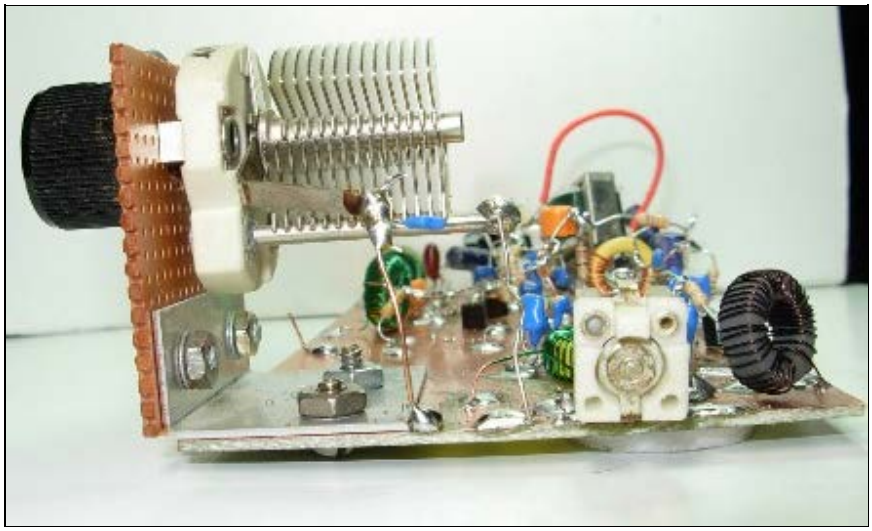
One of the full wave rectifier, voltage regulator and filter bread boards used in these experiments. I went as high as 24 VCC on some single supply amps I tested and was getting over 5 watts average output power


NPN

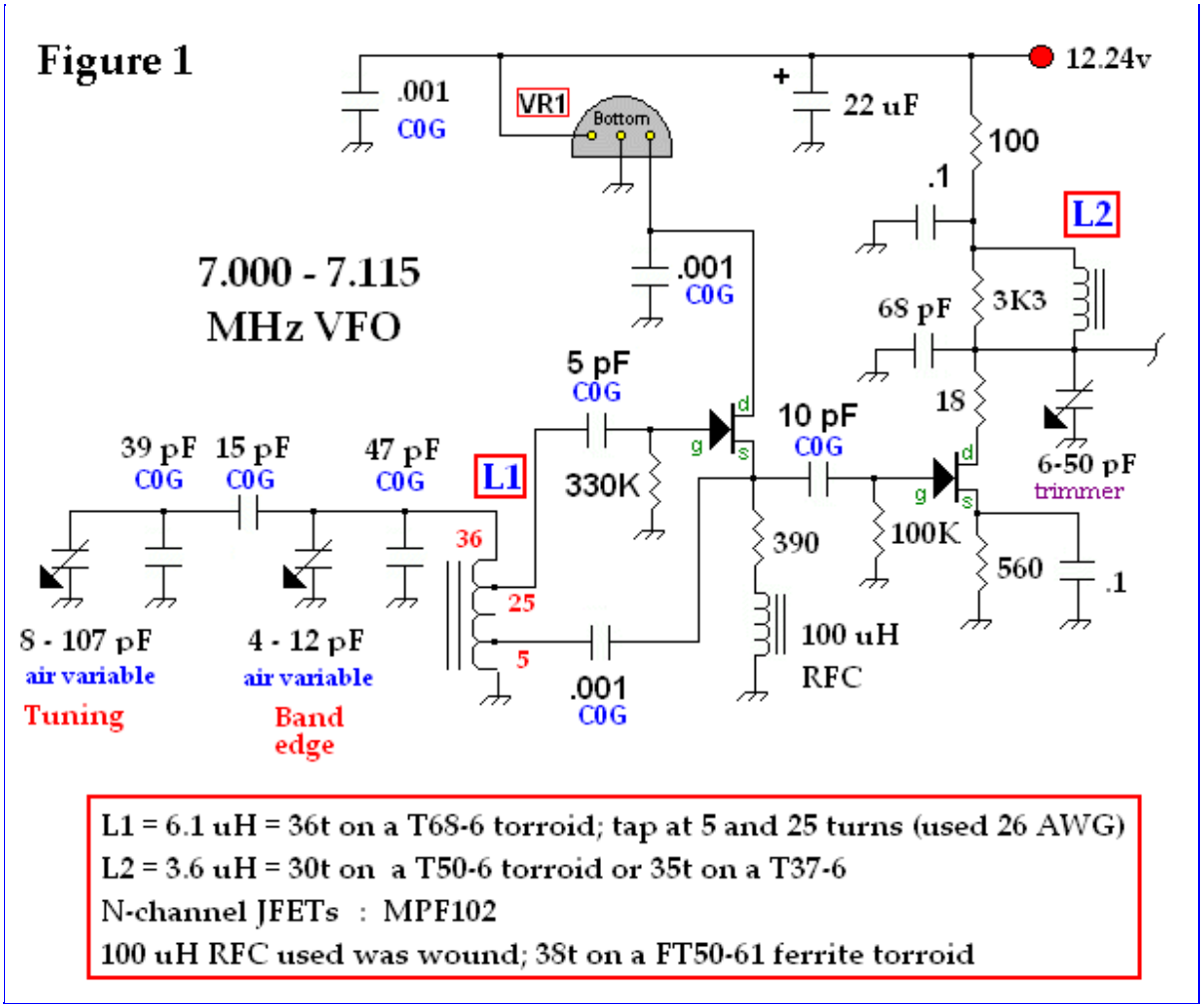
Two Bravo Receiver Experiments

Introduction

An experimental direct conversion receiver is presented. This 1990's style receiver was built to re-familiarize with DC receivers and try out a few new ideas. Design-on-the-bench bread boarding was used exclusively and was a pleasant way to both learn and pass time. Feedback has been received stating that that certain stages of previous receiver experiments were either too basic or too complex and thus a particular receiver was not built. This web site is as much a cookbook as anything. Kludge together whatever receiver stages you want; no project is meant to be set in stone. This receiver has a high popcorn factor with MPF102 and 2N3904s as the main semiconductors.

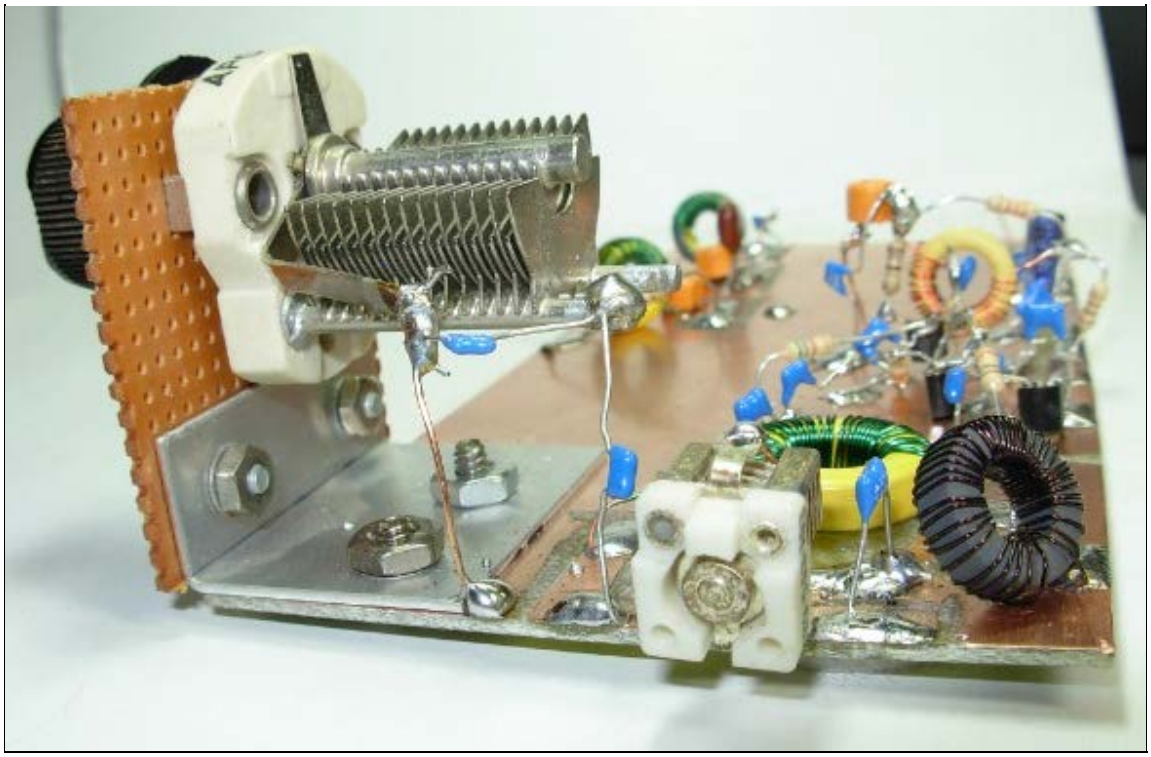


Variable Frequency Oscillator



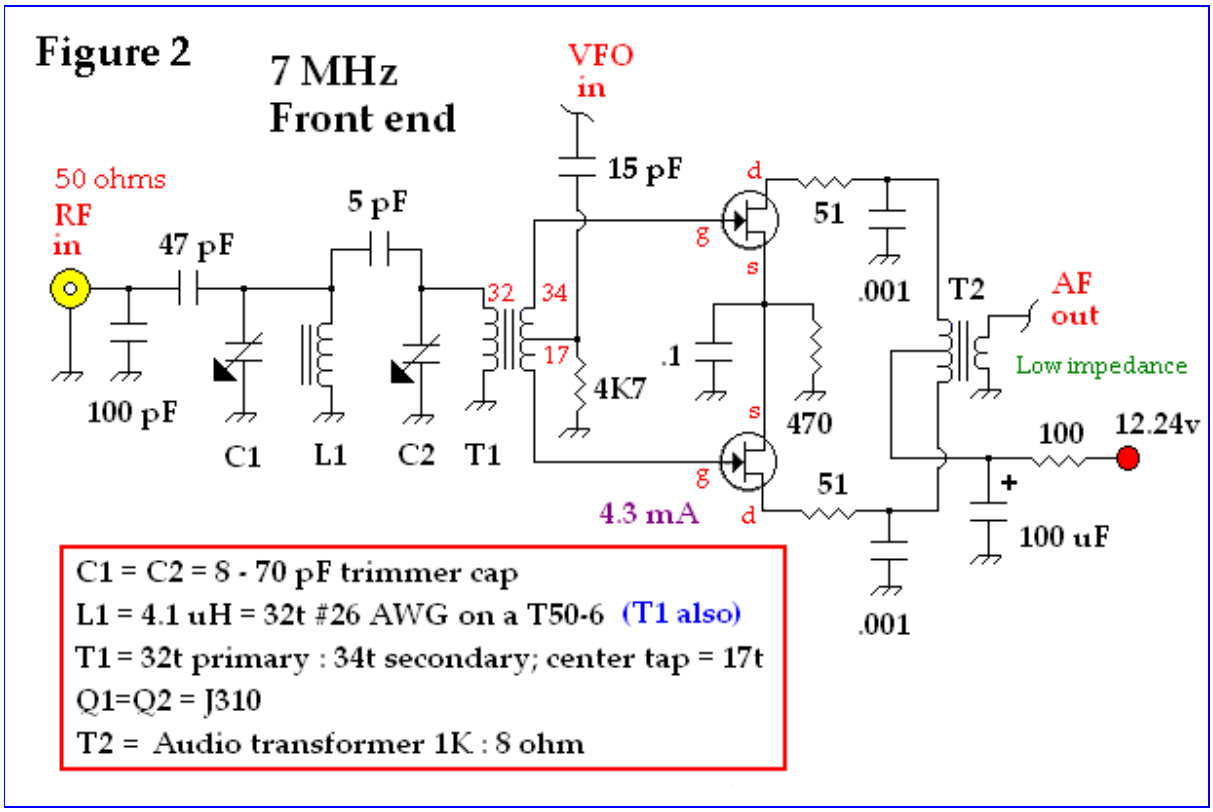
The first stage built was the VFO shown above in Figure 1. The oscillator portion is based upon Figure 4.15 from EMRFD. The VFO resonator tank is isolated from the JFET by tapping down as shown. This is an outstanding VFO topology. See this web page for a few more details and a coil tap calculator. I favor high L to C ratios in my RF tanks, although this does not affect the VFO function. The tapped inductor in this oscillator allows you to use a high RF voltage (low C + high L) while still keeping the FET gate AC voltage at a reasonable level. The buffer amp was designed for high output power and supplies nearly 5 volts peak to peak to the product detector local oscillator port. You can vary the output voltage by increasing or decreasing the 15 pF coupling capacitor for use in other projects.
To peak the L2 tank trimmer capacitor, use a scope, RF voltmeter, or temporarily connect a 10K (or greater value) resistor load to ground via a 10 - 47 pF output capacitor and adjust this capacitor while listening with a nearby CW receiver. (Use a short piece of wire as an antenna.) Additionally, you could also peak this trimmer cap while listening to a CW signal with the completed 2 Bravo receiver. It takes around 100 pF to resonate the L2 tank at 7.040 MHz in case you are wondering.

Since air variable capacitors were used for tuning and to set the band edge, Q is high and frequency stability is excellent. My 1 hour frequency drift was 50 Hertz uncovered. The high RF energy in the tank circuit results in low noise. The L1 taps also allow the use of a 5 pF gate coupling capacitor rather than the hard to locate 3.3 pF cap used in many example VFO schematics. With different buffer/amps as required, this is now my number 1 VFO topology and it is nothing short of stellar. Note that the 100 uH RFC can be wound with 15 turns on an FT37-43 ferrite torroid, or replaced with a fixed value choke.



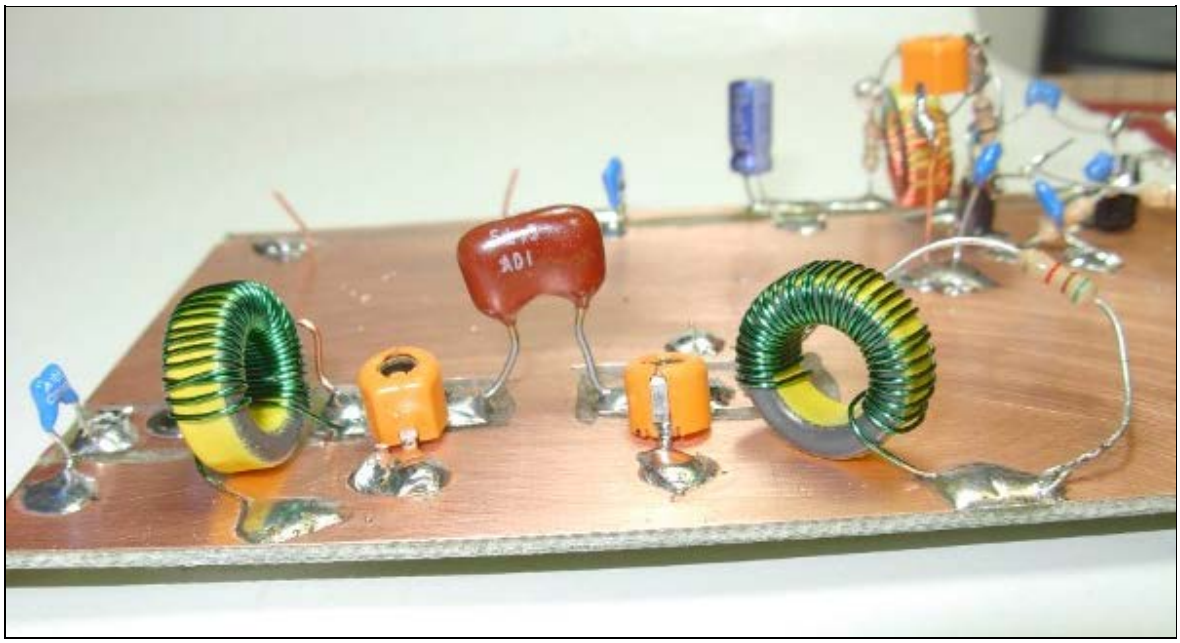
In the above photograph is the VFO bread board. I used 26 gauge wire for the inductor and took my time to make sure the wire was laying flat on the T68-6 torroidal core. You can pull the wire tighter if you wash your hands before winding.

Band pass Filter and Product Detector



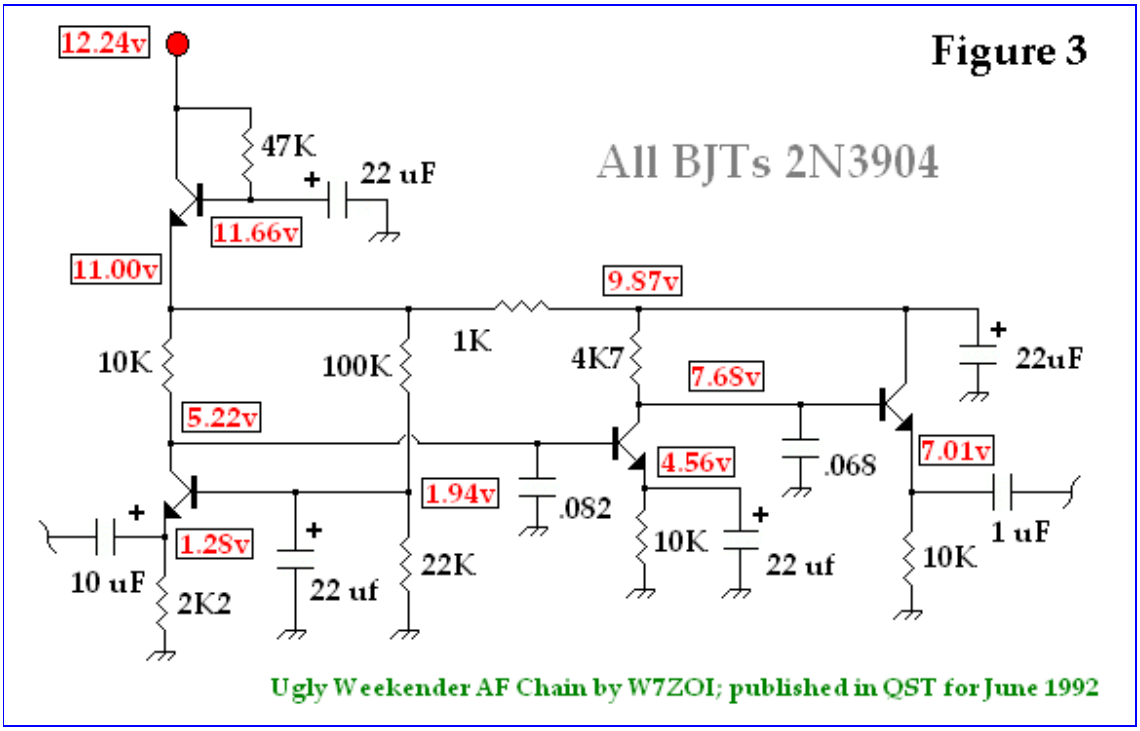
Please refer to Figure 2. The second stage constructed was the double tuned band pass filter. You will need about 50-54 pF to resonate L1 and T1 at 7.040 MHz. The C1 and C2 values chosen are thus perfect for tuning the 40 Meter CW band. I peaked my particular front end filter at a center frequency of 7.025 MHz using a 50 ohm output impedance RF generator and then did some fine tuning with an antenna connected after the receiver was constructed. You may also just tune C1 and C2 for maximum signal strength when listening to band noise and QSOs. Filter bandwidth is sufficient to cover the whole CW sub-band. No AM broadcast band radio was heard during several nights of testing.

The product detector is single balanced for improved port isolation and BCB rejection. Lay out your circuit to try to achieve symmetry. The schematic calls for J310s. I built the first prototype with MPF102 that were matched for Idss. To find two with the same Idss, I had to measure 16 transistors! This is too painful, and I recommend just using a pair of J310s. The words "matched" and "MPF102" should not be used in the same sentence! Ideally, your J310s should be matched, however, the process should not take as long as for MPF102 JFETS. The choice is yours to make. T1 is a little tricky to wind, however, your best effort should be good enough.
Some builders will be unhappy with using a audio transformer (T2), however, they are still in catalogs and online stores, or can be harvested from an old transistor radio. CB radio modulation transformers are also a possible source. A higher impedance audio transformer, will likely give even more conversion gain. Without the 51 ohm drain resistors, oscillations occurred in my bread board.



Above. A temporary 5K1 (5.1K) resistor was soldered across the second tank for testing when the front end filter was designed on my work bench.

Audio Pre-amplifier



To match the low impedance winding of the audio transformer, a common base amp topology was chosen. I decided to use a favorite circuit; the

audio chain from the first amateur band receiver that I ever built -The Ugly Weekender. The final common emitter feedback amplifier from the original schematic was omitted as the 3 stages above provided enough voltage gain. The Figure 3 amplifier is worth studying. It is difficult to DC couple audio amplifier stages and not end up with your second and/or third stage in saturation. This example of good design by W7ZOI illustrates how to do it. The third stage, a common emitter amp is a level shifter and drops the DC voltage back down, although this stage is AC coupled to the volume potentiometer. For lower noise, you could AC couple the first common base amp to a 5532 op-amp, although, this would reduce the popcorn factor a bit. Do not expect ear blasting voltage gain from this humble circuit. It provides reasonable drive to the power amp stage.
The 0.82 and 0.68 uF capacitors shunt any detected RF energy to ground and also provide some low pass filtering. The original schematic called for 0.1 uF capacitors and which values work the best is yours to decide. Polyester film type capacitors were used in the bread board.

Photographed above is the Figure 3 bread board. You can also see the VFO buffer/amp and the audio transformer. I tried several AF preamplifiers, but preferred Figure 3 to all others.

Audio Power Amplifier

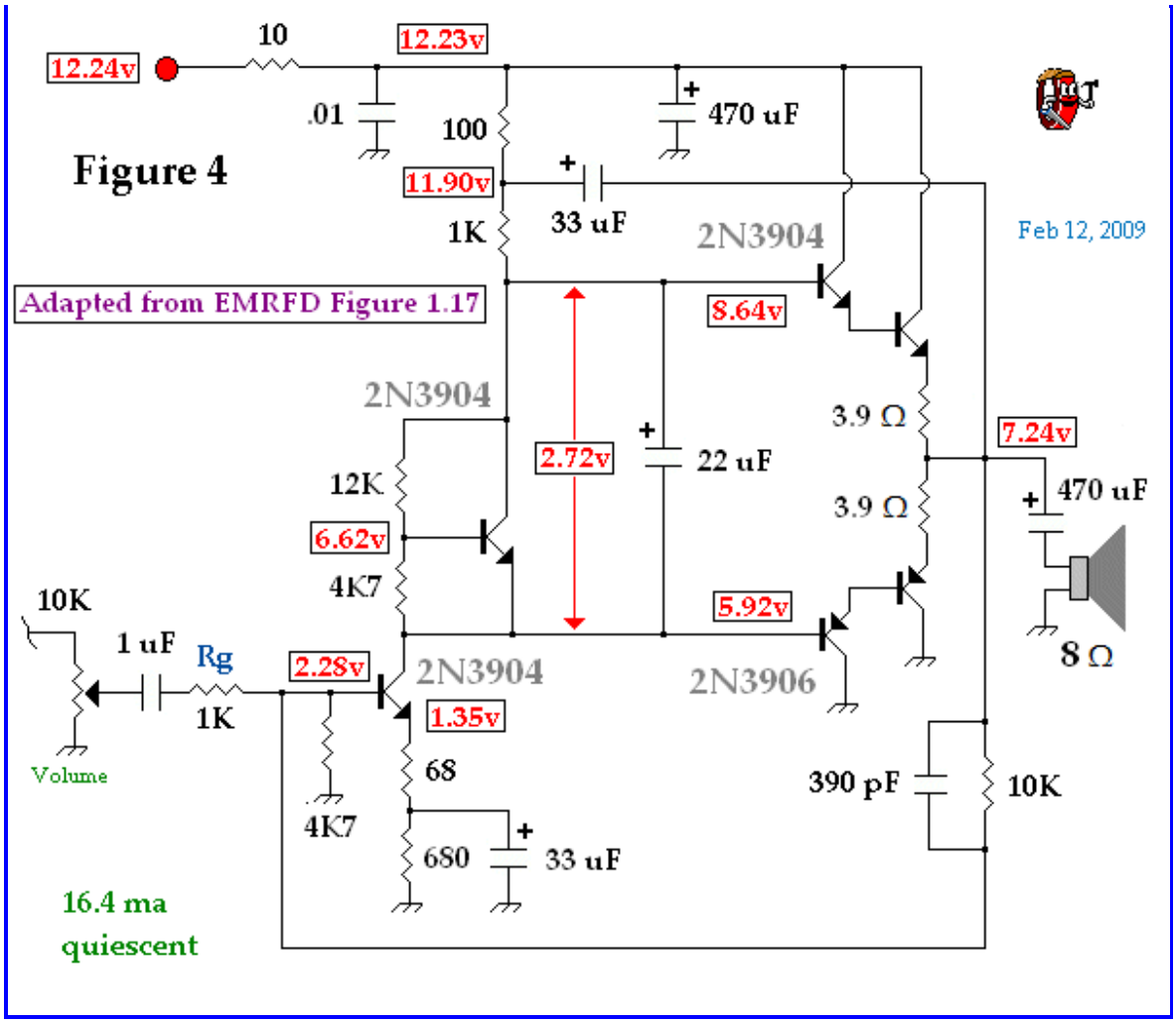
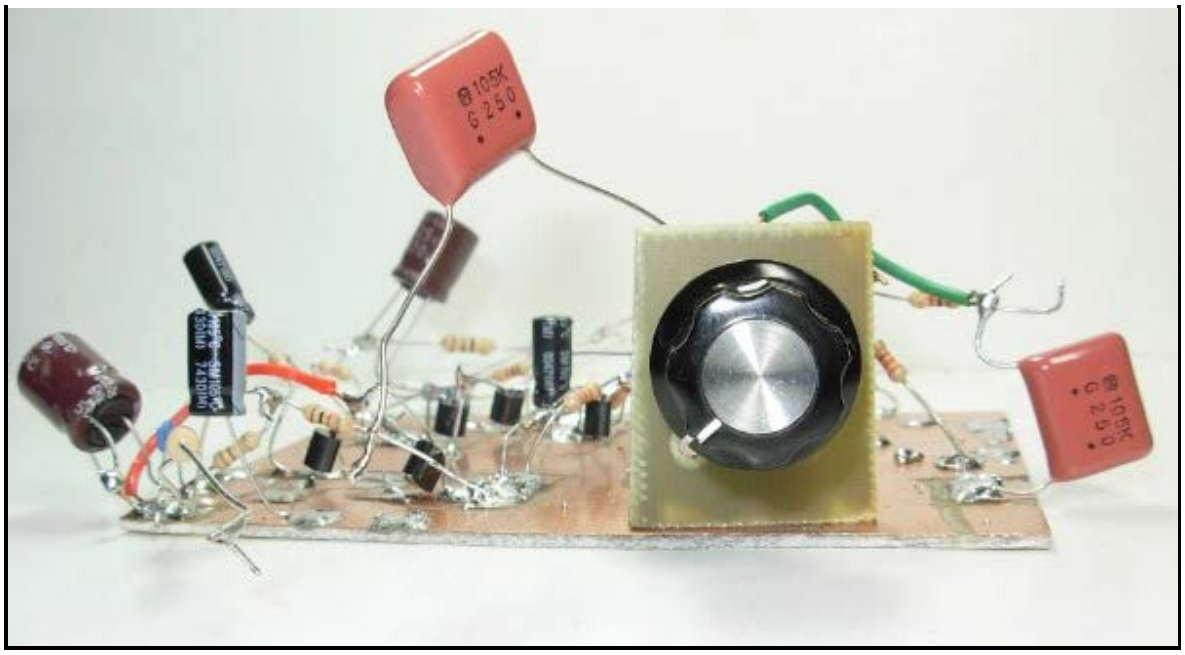


Figure 4 provided 3 nights of experimentation. The base circuit for this amplifier was Figure 1.17 from EMRFD. To increase power gain and reduce harmonic distortion, Darlington configured emitter follower pairs were employed. This worked, except the power followers were under biased and had serious crossover distortion. To remedy this, the amplified diode (level shifter) bias was increased until a sine wave was seen on the oscilloscope. This was achieved by replacing the 10K resistor (R6 in the original schematic) with a 4K7 ohm resistor. The next task was to try to increase the voltage gain. Rg in EMRFD Figure 1.17 called for a 3K3 resistor. Rg was dropped to 1K; this worked. The degenerative feedback on the common emitter amplifier of EMRFD Fig 1.17 was also dropped somewhat. Next some bootstrapping feedback was AC coupled to the collector of the main common emitter amplifier. Each of the 2 collector resistor values was changed around and the outcome was recorded. Ultimately the 100 ohm plus 1K ohm resistor series pair was chosen and provided a boost of 0.85 volts peak-to-peak clean voltage gain to the output waveform. The power follower emitter degeneration resistors were also decreased from 22 to 3.9 ohms. The result is a low distortion power amp with about 150 mW of clean average power output. This receiver is not super loud, but it is reasonably loud and the audio is bell clear. If you use this amplifier stage in other projects that have a higher gain pre-amplifier, I recommend keeping Rg at 3K3 ohms as this amplifier will likely exhibit lower distortion characteristics.



Shown above is the Figure 4 bread board. A 10 ohm resistor was used to decouple this stage. Without the resistor, audio oscillations at around 850 Hz manifested when the volume was greater than about half way up. The voltage drop across the 10 ohm resistor is trivial. You may have to increase this resistor value if you experience instability. Expect all amplifiers to oscillate and decouple them accordingly. The 390 pF feedback capacitor is required. The Figure 4 amplifier exhibits greater gain as frequency increases and in a direct conversion filter with no low pass filtering, this would be very harsh indeed. Feel free to experiment with the value of this feedback capacitor. Kudos to W7ZOI for the EMRFD Figure 1.17 schematic which serves as a great specimen to inform and challenge us experimenters. The original common emitter amplifier (Q1 in EMRFD Figure 1 .17) bias current is perfect and although I tried increasing and decreasing it, I returned to the originally specified bias resistor values.

Additional Outputs

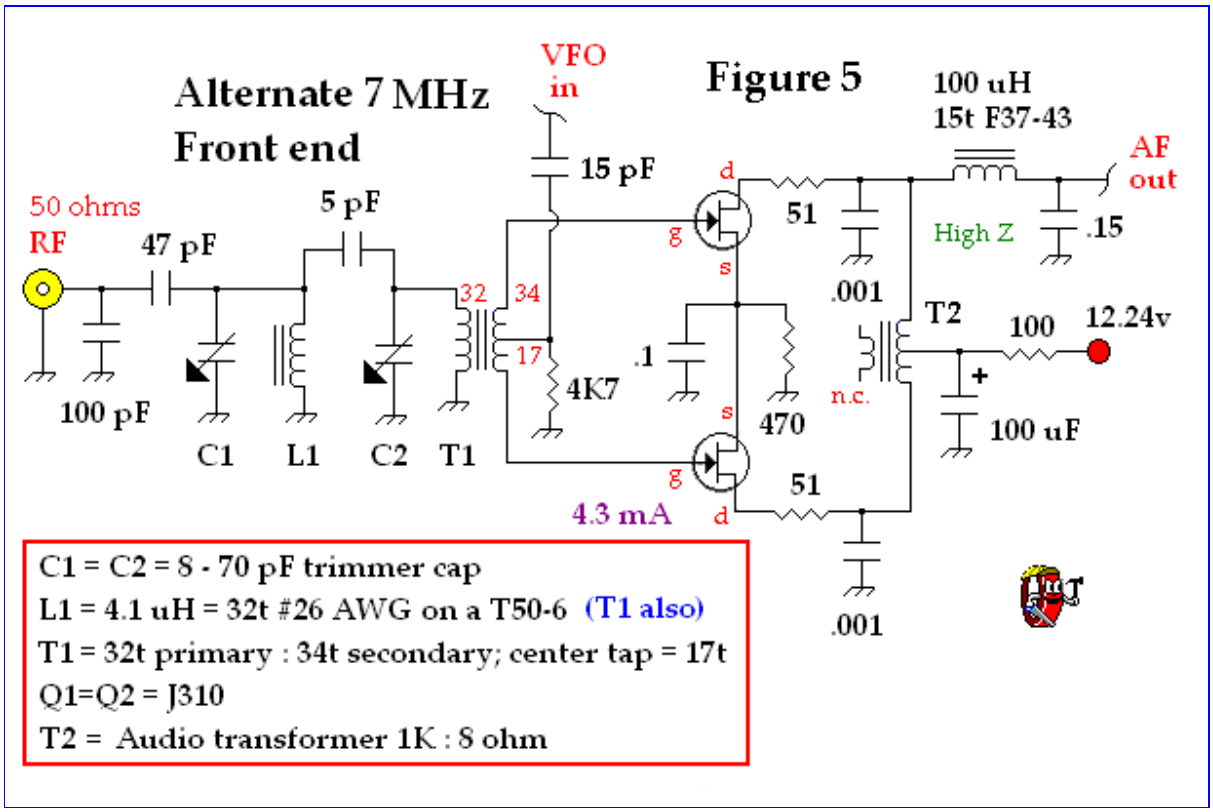


Figure 5 depicts the product detector with a high impedance output. From my experiments at least, it was better to use the low Z coil for improved product detector balance and audio voltage gain.

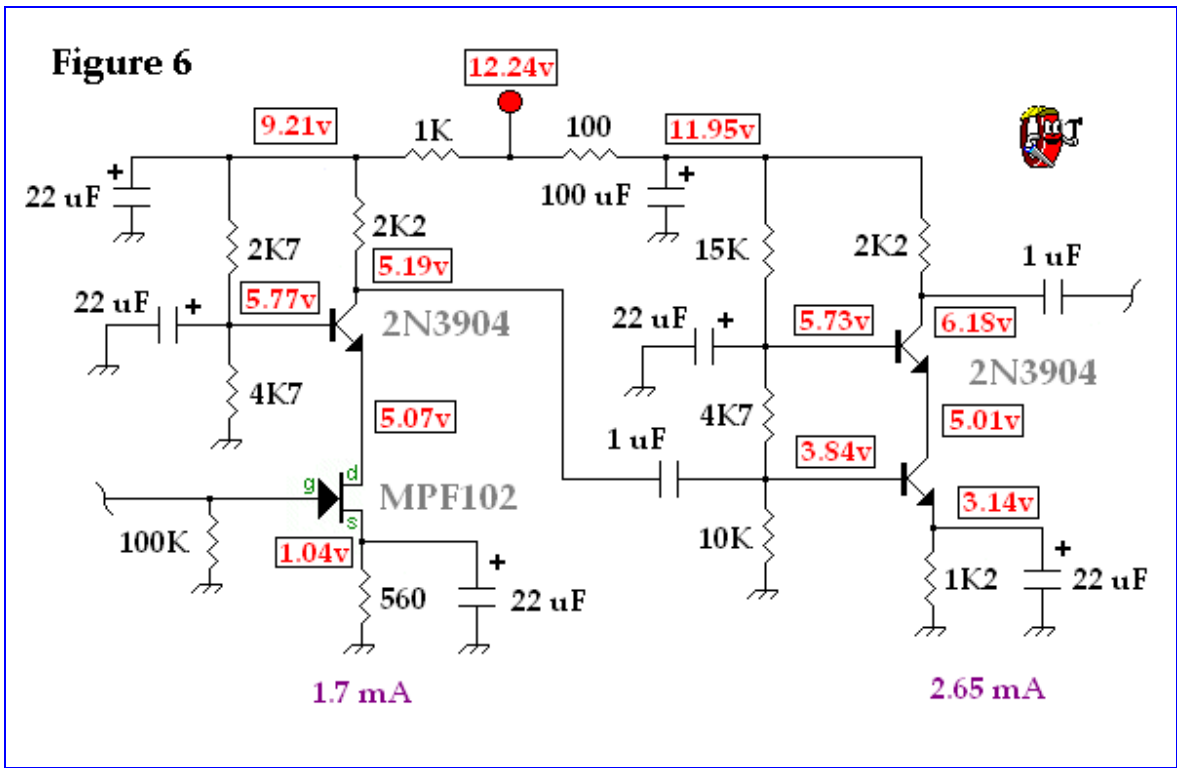
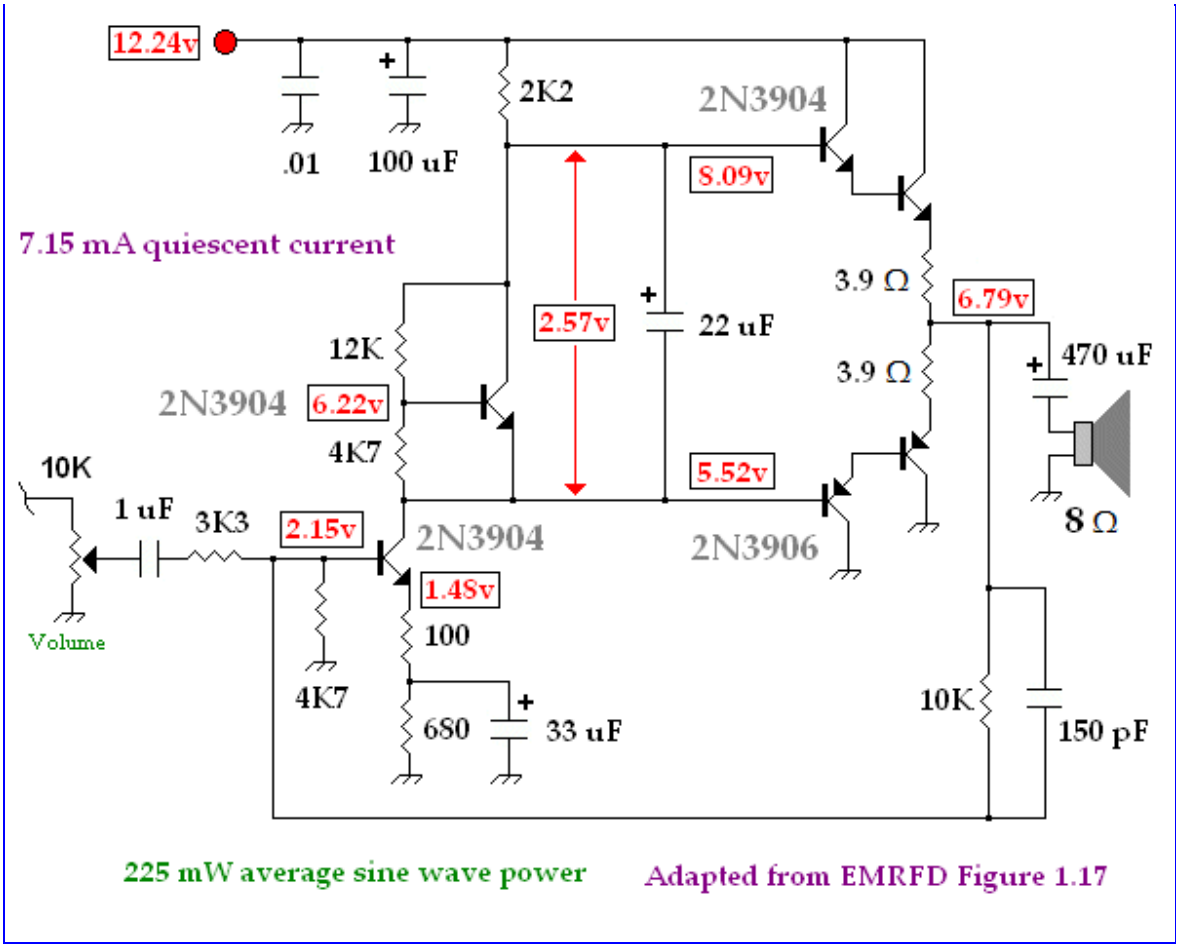
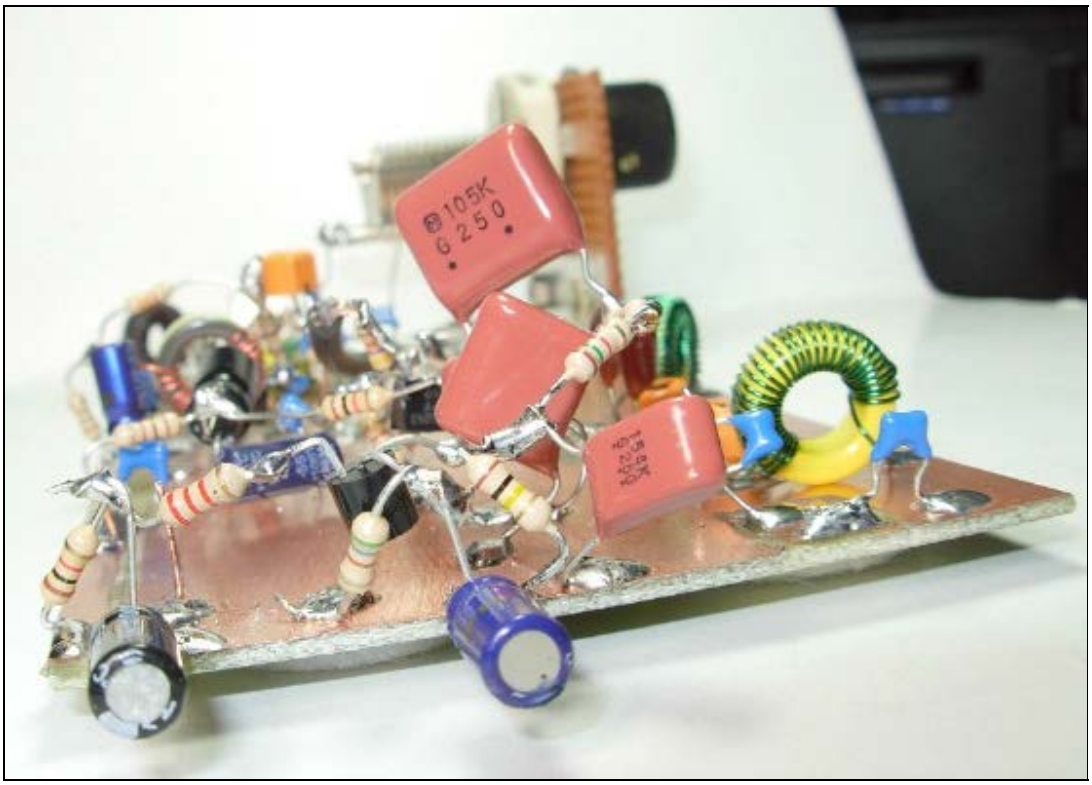


Figure 6: A high impedance input audio stage. The first stage is a hybrid cascode. The second stage is common emitter, common base cascode. Care was taken with transistor biasing to try to optimize distortion characteristics. Certainly, I am a total novice with such amplifiers and more time on the bench and also with computer simulation is required to better understand these amplifiers. At any rate, the schematics with DC voltages are posted for others to study and hopefully improve.

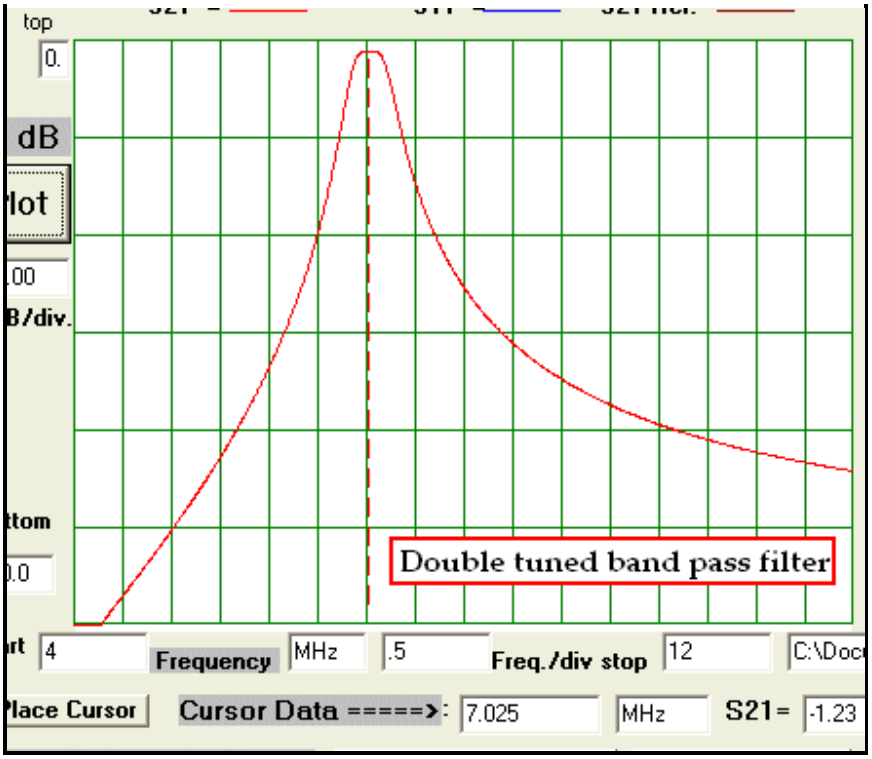


I have 3 adapted versions of the EMRFD Figure 1.17 amplifier in my note book. This is my favorite and has the greatest clean maximum average power output of all of the 3 versions. This power amplifier somewhat lacks sufficient voltage gain for the 2 Bravo receiver (with its relatively low gain audio pre-amplifier) and thus Figure 4 was chosen as the more suitable power amplifier. You could lower the 3K3 resistor to increase the voltage gain. This circuit begs experimentation.



A couple of low pass RC audio filters were tried but later abandoned. One filter with its 1 uF AC coupling capacitors is shown in the photo above. It is quite an experience to hear an unfiltered direct conversion receiver. I love the purity. This is okay for an experimental or casual receiver, but not for a contest rig. Low pass filtering is definitely required in that context.

A different angle photo of the VFO bread board. My "build most of the project on 1 copper clad board" construction technique is not really suitable for "a keeper" receiver. VFOs should ideally be in a shielded box. Proper construction techniques and grounding ideas for DC receivers can be found in EMRFD, so they are not covered on this web page.



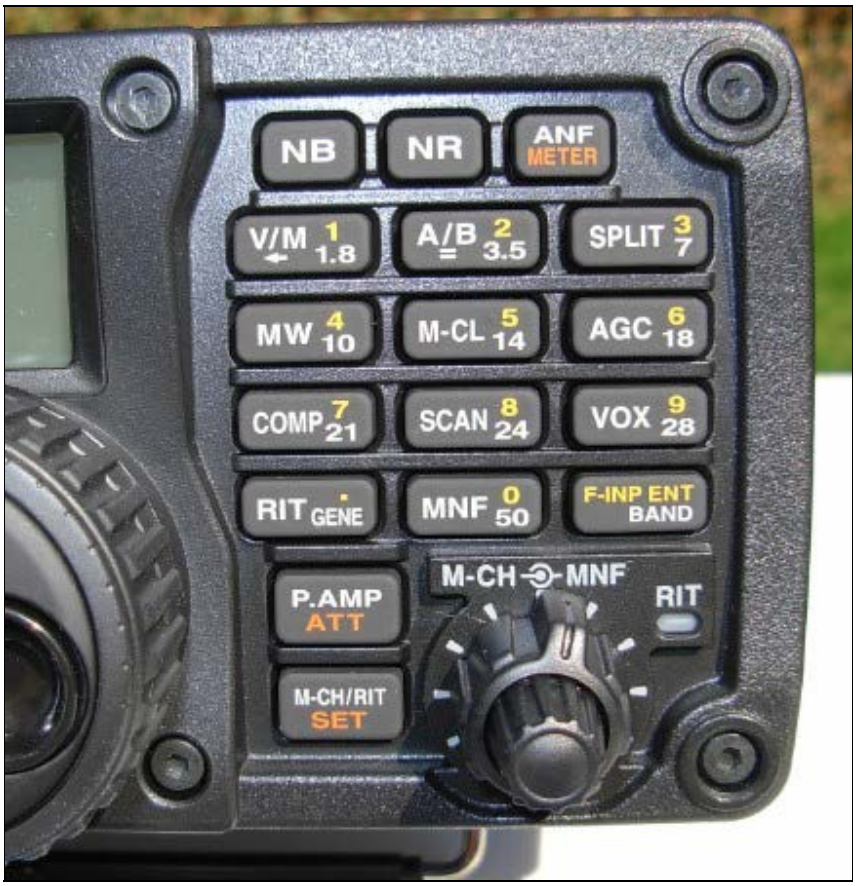
A GPLA simulation of the front end band pass filter centered at 7.025 MHz.

Amateur and Short Wave Radio Electronics Experimenter's Web Site

ICOM IC-7200 Listening Tests and Observations

Introduction

In Spring 2009, I evaluated and photographed the ICOM IC-7200 transceiver. For this web page, only its receiver was evaluated for HAM and SWL purposes by using my eyes and ears. For technical evaluation, please click on these assessments written by Adam, VA7OJ/AB4OJ or Peter, G3SJX . Additionally, eHam.net has a review web page to consult and the ARRL publication, QST for June 2009 has a review. With my modest camera and lighting equipment, it was difficult to well photograph this very dark colored transceiver indoors.



I like ICOM radio equipment and am therefore biased in my review. Please consider trying this and any other radio out before you purchase it. Our needs, expectations and budgets tend to be uniquely different. Like others, I would rather own the new ICOM IC-7600, Yaesu FT2000D or Elecraft K3, however, my budget does not allow this.

It appears that the target audience for this transceiver is as follows: portable or field/emergency communication usage and/or it is oriented towards entry level HAMs or perhaps those wishing a modern DSP-filtered back up rig. My review is from a SWL/HAM perspective.

General Thoughts

Although modern and stylish, this radio is very easy to use. The owners manual is well written with clear examples of how to perform the various setting changes. After reading the manual and and trying the radio buttons and knobs out, I pretty much mastered receive operation on my first night. The LCD display is small, however is reasonably adequate considering that this whole radio is compact. I believe some operators will have difficulty with this diminutive display. The S-meter is a little difficult to see. You can push a front panel button which announces the S-Meter reading, set frequency and mode in English or Japanese language if you can't quite see the display at an odd viewing angle.

The VFO knob has adjustable resolution and works well, however, it lacks that silky smooth/weighted feeling of many other radios including the R75 receiver. The ICOM engineers had there work cut out for them; include all the modern interference fighting features; place them in a small chassis; make it easy to use and come in on budget. They did it! This radio is full of useful interference management tools. One example is the (DSP) manual notch filter; it works superbly. There is no FM mode. I imagine by dropping FM mode capability , the designers were able to use the specified 6 KHz roofing filter at the second IF which has the potential to improve dynamic range at close-in signal spacing for some modes. When I started in Amateur Radio many years ago, roofing filters were never mentioned, but have become a huge marketing lever and seemingly a topic of much confusion.

DSP IF Filtering

The DSP IF filtering works very well. Does it function better than more traditional crystal IF filtering? Yes and no. It is a question of compromise for me. The greatest DSP attribute is that there are no expensive crystal filters to purchase. Additionally, you can customize the desired IF filter bandwidth (wide, medium and narrow) for each mode and also set a "hard" versus "soft" filter shape. The soft shape equates less ringing and potentially less listener fatigue than the hard setting. I have never liked listening to CW signals through stiff, 6-8 pole 250-500 Hz crystal filters on any receiver, so for me, adjustable DSP IF filtering is preferable. DSP IF filtering is not perfect as authors like Rob Sherwood, NC0B have presented, however, ICOM have a pretty good DSP platform out now and I am pretty sure their latest technology from the IC-7600/PRO 3 on down have also been used in this receiver. After all, you can always try to find an old Drake and order some crystal filters for it, or spend significantly more cash on a high-end transceiver if you need better performance.

Shown to the the left is a few front panel controls and the front firing speaker. This speaker is reasonably nice sounding. There seems to be less harmonic distortion and audio gain than the R75 receiver. The radio front panel controls are well thought out and seemed intuitive to use after reading the manual and/or just trying them out. The various hardware components such as potentiometers, rotary switches, microphone jack, PL-259 jack etc seem not to be of the highest quality. When you compare such components to that used in their older designs, there is

evidence of modern cost containment. I have seen many reviews describing this transceiver as rugged. Certainly the diecast frame is solid, and I hope the forementioned hardware is as well. There is only one antenna connection. I believe that any radio offering the 6 meter band should offer two SO239 antenna connections, however, this likely would have crowded the back panel layout and increased cost for the RF in/out switching.

The VFO knob (shown to the right) looks and feels a little cheap. Functionally, however, the knob spins very well and precise tuning is easy to perform. One of my first tasks was to input various the frequencies I use for both amateur and short wave listening into the (201 maximum) memory slots. This was very easy to do and each memory channel also stores the mode and filter setting.

AM Reception(R75 versus IC-7200)

The IC-7200 is a good short wave receiver. With a high quality external speaker connected, pleasant and warm sounding AM audio may be heard; however it is not Hi-Fi. That is; the IC-7200 is not an audiophile AM receiver. There is no synchronous AM detection and no 15 KHz wide IF filter for example. For AM reception, the user may choose from an 8000 to 800 Hz IF bandwidth in the 3 switchable filter settings. During testing, I set my wide filter setting to 8000 Hz, my medium setting to 6000 Hz and my narrow setting to 3000 Hz for AM reception. Of course these bandwidths can be further adjusted at any time. Local MW broadcast radio sounded great with an 8000 Hz IF bandwidth during testing. The various AM reception appropriate interference controls such as pass band tuning (PBT), automatic notch filter (ANF) and digital noise reduction are all configurable or adjustable and I found the ANF and digital noise reduction helpful when listening on the crowded 49 meter band during a rain storm.
I performed A/B comparisons with the IC-7200 and the R75 simultaneously hooked to the same antenna. For AM reception, they are comparable with the R75 having marginally better sounding (more Hi-Fi) audio. I had the R75 IF filter bandwidth set at 15 KHz and the IC-7200 IF bandwidth set at 8 KHz (both at their maximum IF bandwidth).

CW/SSB Reception (R75 versus IC-7200)

In CW mode, both were set to have a 500 Hertz IF bandwidth.(The R75 had the FL-100 model 500 Hertz crystal filter in its 9 MHz IF slot). The IC-7200 was quieter, clearer and less overloaded by adjacent stations during pile ups. There were occasionally weak signals I was able to copy on the IC-7200 that I could not even hear on the R75 due to noise. The IC-7200 is a joy to use on CW; really fantastic.
On SSB, I found both receivers fairly comparable, but the IC-7200 was better for pulling weak signals out of the noise as the noise floor was a little quieter and the audio a little more crisp. My wife also agreed with me in her "independent tests" of the CW and SSB reception. I am certain to catch flak because my subjective comments are based upon listening tests and not measurements. It is interesting to note that in almost every aspect of our lives, preferences are made using 1 or more of our 5 senses. From listening, observing and also reading the reviews of others, it seems ICOM has a hit with the IC-7200.

Mouse over the images below left to view a full size photograph

The IC-7200 in the sun.

Rear view of the transceiver.

Top angle view. There are ventilation holes at the top right hand side. You can just see the ventilation screen just next to the right top panel screw.

Reverse angle top view. The military look is attractive.

The IC-7200 and R75 side by side.

Another photograph of the 2 radios from the front angle for comparison. The LCD display size difference is quite obvious in this shot.

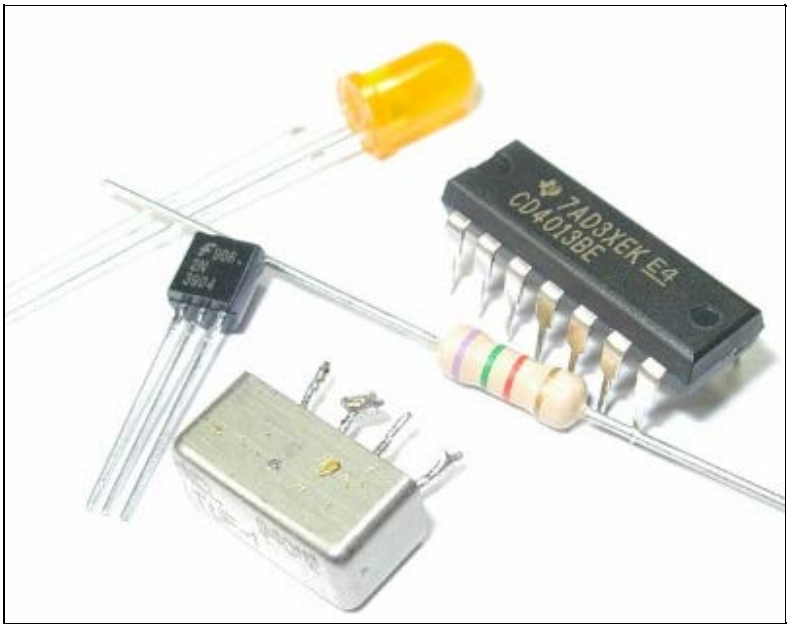
The receivers at a right angle. The IC-7200 is a glad update, although, I would likely not upgrade from the R75 to the IC-7200 just for receiving purposes. If I was considering choosing the IC-718 (transceiver version of the R75), the IC-7200 is worth the extra money. It is also an excellent transceiver consideration for an entry-level rig, for back up purposes. or for the budget-minded operator.

Conclusion

The IC-7200 is a serious choice for amateurs seeking a good, modern HF plus 6 Meter transceiver. It is more than adequate as an SWL receiver if you are an amateur radio operator plus SW listener like myself. Please try one out for yourself if you are considering this transceiver.

Amateur Radio Electronic Design

Electronic Hobbyist Circuits



This page will house a collection of brief hobbyist experiments.

1. Pseudo-Random Number Generator

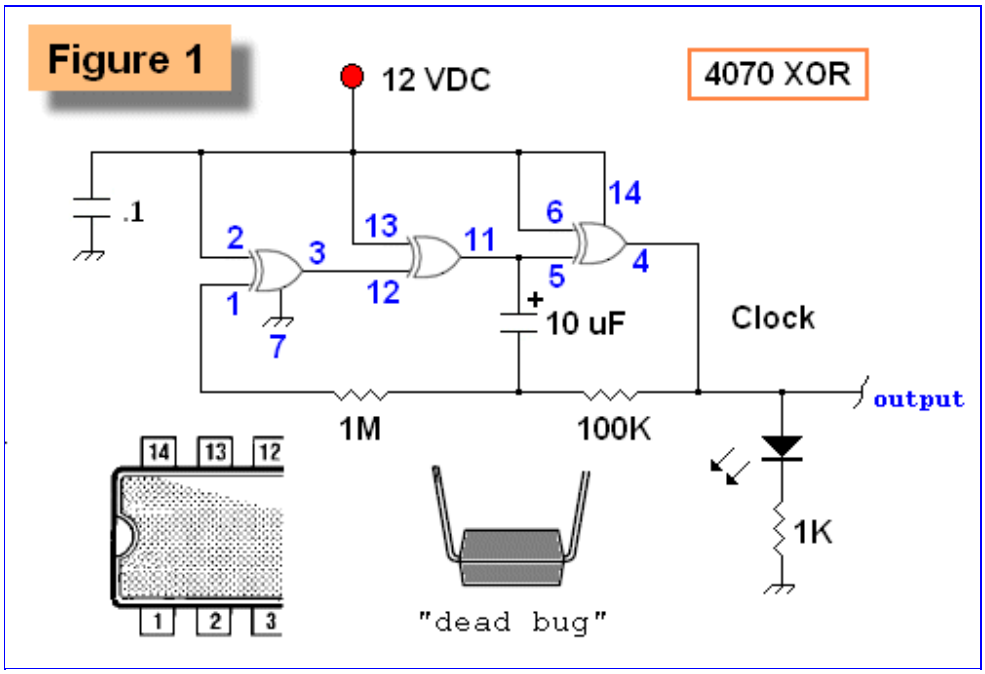
This circuit describes a simple, 6-bit random number pseudo-generator used to study binary counters and in particular, shift registers. Some very basic background information about binary counters and shift registers is provided. In reality there are dozens of different shift register topologies available and it can get quite complex. If you wish to find a good logic tutorial website, I strongly recommend Ken Bigelow's site as it has interactive diagrams. Flip-flops are also covered well on wikipedia and many other web sites.

Binary Counter: The circuit most often used as a counter is called a binary flip-flop. The basic flip-flop can be viewed as a toggle switch having either an ON or OFF position. This is the binary state 1 (HIGH) or 0 (LOW). Like the toggle switch, the binary flip-flop has 2 binary states 1 or 0. A binary flip flop counter counts in a sequence such as 0, 1, 0 ,1 etc. A straight binary counter can be built by using 1 or more flip-flops connected in a manner that the binary number stored in these flip-flops will represent the total number of trigger pulses received at the counter input.

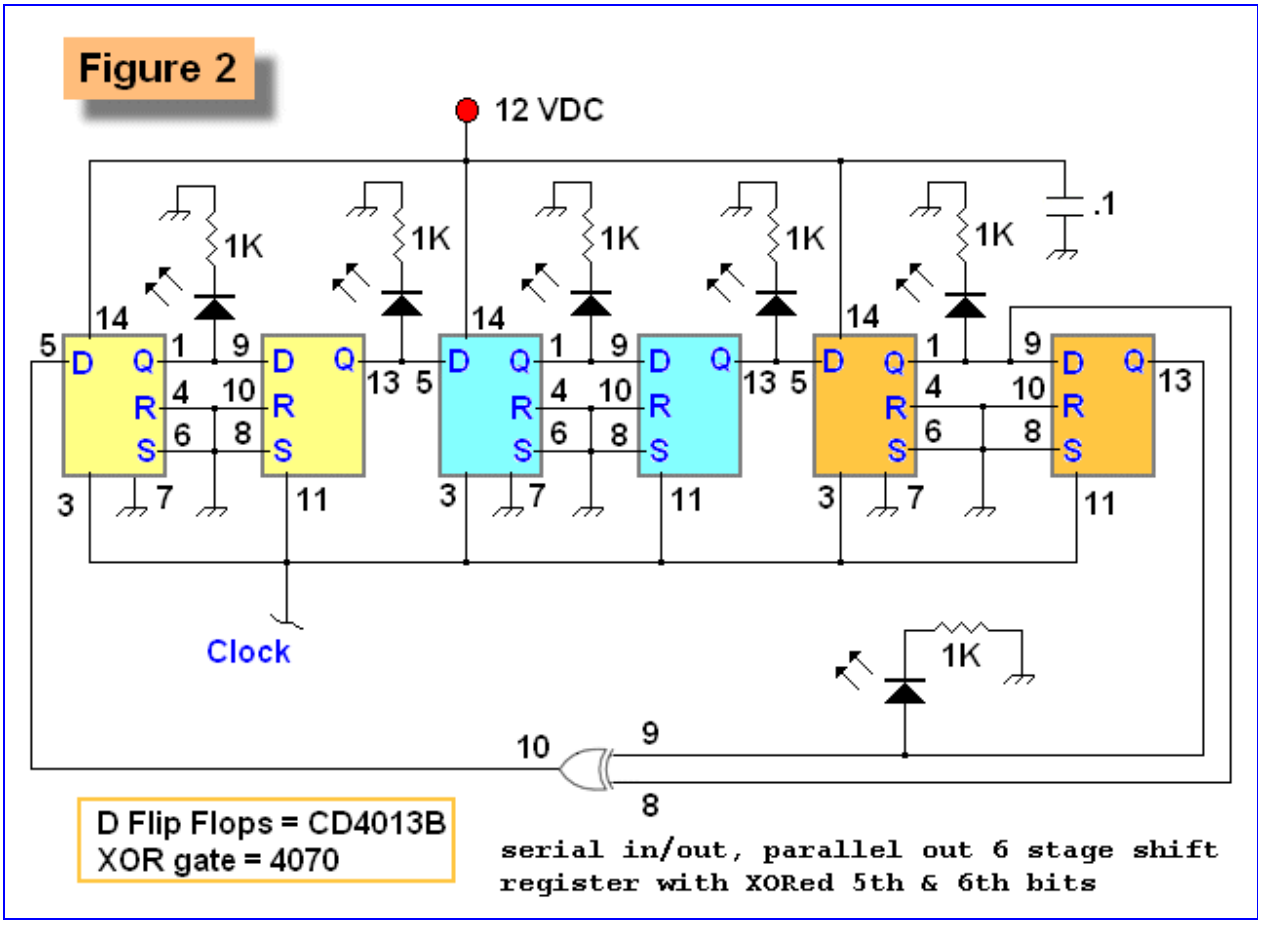
Ring Counter: A ring counter has 2 or more flip-flops cascaded so that the output from one flip-flop becomes the input of the next flip-flop. The flip-flops are connected so that all of their outputs are at the binary state 0 except for one flip-flop. By pulsing the input of the ring counter, it will sequentially change the binary state of the succeeding flip-flop from binary 0 to binary 1. The flip-flop that contains the binary 1 indicates the count of this binary counter. The maximum number of pulses that can be counted by N flip-flops is N pulses.

Shift Register: A serial entry shift register is similar to a ring counter, except that the output flip-flop is not connected to the input flip-flop. Like the ring counter, the flip-flops are cascaded so that the output from one flip-flop becomes the input of the next flip-flop. All the set trigger and reset trigger inputs are tied together to form what is called a shift bus. Clock pulses are applied to the shift bus to cause the stored binary information to shift from left to right; one bit position per each received clock pulse. In Figure 2, this serial input/output + parallel output register has its 5th and 6th bits exclusive ORed to the serial input to form a pseudo-random sequencer, which is called a pseudo-random number generator by some.

The CMOS logic ICs used were one 4070 XOR (Exclusive OR) and three CD4013B D flip-flops. Junk box LEDs were used to observe the binary state of the clock and each of the 6 bits of the shift register.

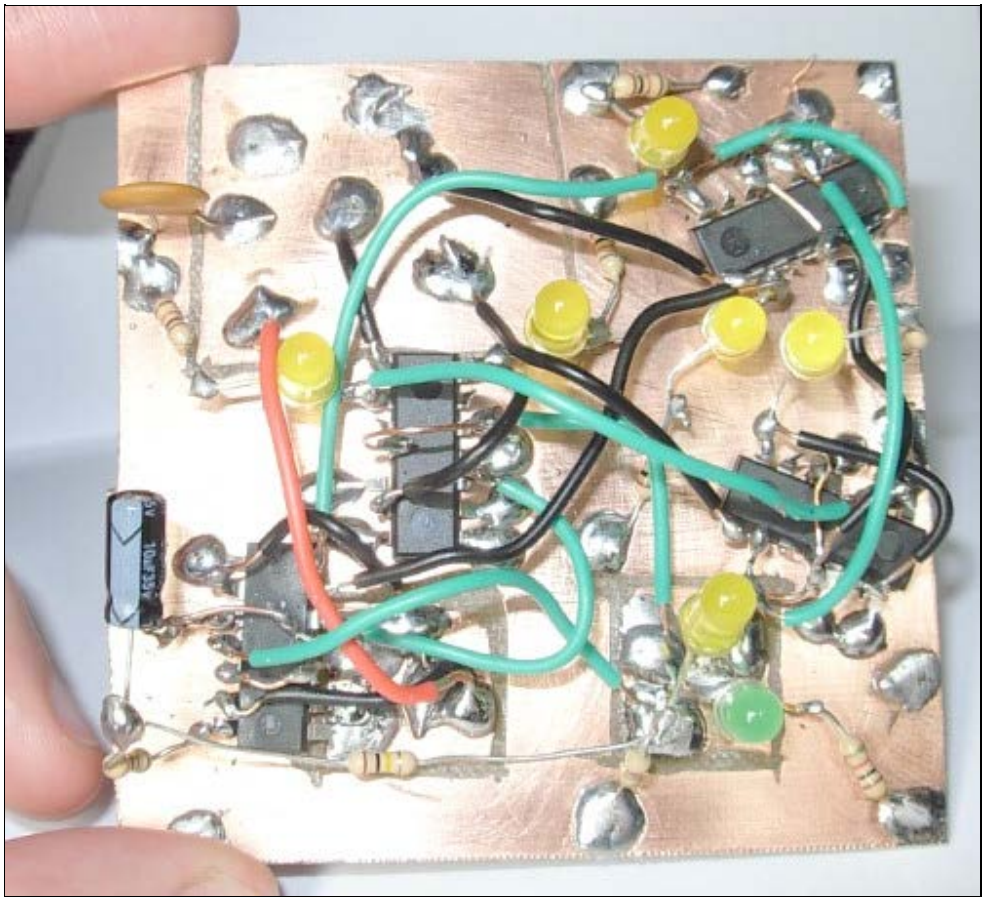


Since only one XOR gate is needed for the shift register, the remaining gates were configured to make the clock. These gates are essentially wired up as inverters to form an astable multivibrator with a frequency of about 0.45 Hertz or 27 pulses per minute. Shown above in Figure 1 is the clock schematic and the pin 1 marking for all of the digital ICs on this web page. The output LED is not mandatory, but will instantly tell you whether or not your clock is working. I built this whole circuit using Ugly Construction with the ICs flipped upside down in a "dead bug" fashion. You can increase the clock speed by decreasing the 100K resistor or the capacitor values. F Hertz = 1/ (2.2 * R * C) with R in ohms and C in farads. The slow clock speed was chosen to better observe the digital output of the shift register.



In Figure 2 is the shift register. Each 4013 was wired up as 2 cascaded flip-flops and connected to the clock. Power was applied and then a test lead was used to bring pin 5 of the first flip-flop HIGH (connected to 12 volts for 1-2 seconds) . Both flip-flop state monitor LEDS turned ON in sequence with subsequent clock pulses. Afterwards, pin 5 was set LOW (shorted to ground with a test lead for a couple of seconds) and each LED turned OFF in sequence with subsequent clock pulses. The remaining two 4013s were wired up and tested the the same way and then finally the last XOR gate was wired up.

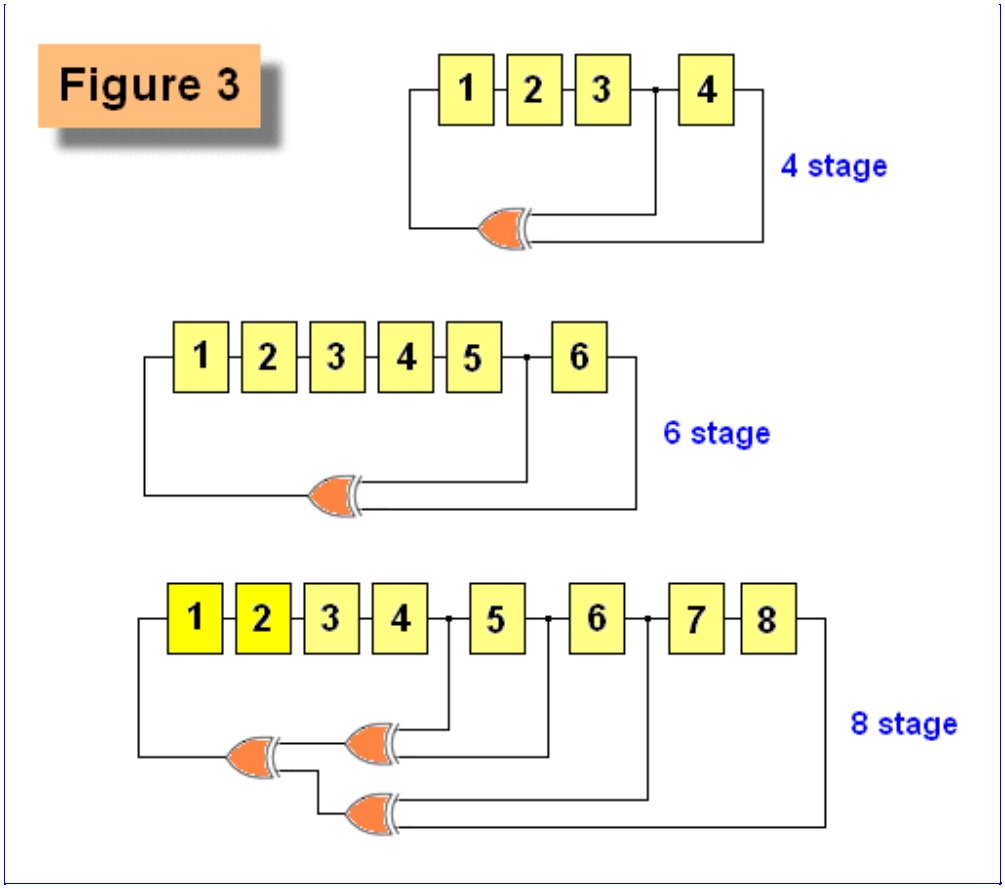
To avoid error, frequent pin counting and a systematic approach is recommended. For example, for each 4013, I soldered the ground pins, wired the pin 14 VDD, connected the clock to pins 3 and 11, then wired up the pin 1 and pin 13 LEDs. Systematic construction techniques are something that you the experimenter can develop and perfect over time. This approach saves time and grief. On some projects, when you have a lot of pins wired up, tracing and repairing an early mistake can be difficult.



Shown above is a bread board of the entire pseudo-random number generator. I just built in on a scrap of board and did not lay it out so the LEDs were in a row, as I am not going to keep this project. The clock state monitor is the green colored LED. There are 63 possible states or combinations of the 6 bits (111000, 100110, 100101, 000101, 000001 etc.) State 000000 is disallowed and will hang up the shift register. If your clock LED is flashing and no shift register LEDs are lit, then "reset" by momentarily setting pin 5 of the first flip-flop HIGH (momentarily apply 12 volts). Long live the reset switch!

Pseudo-random numbers are now mostly generated by computer microprocessors controlled by software and have applications in cryptography, electronic music, security and many other applications. This "hardware" pseudo-random number generator experiment was really cool and if you want to randomly flash some LEDs, this could be the project to use!

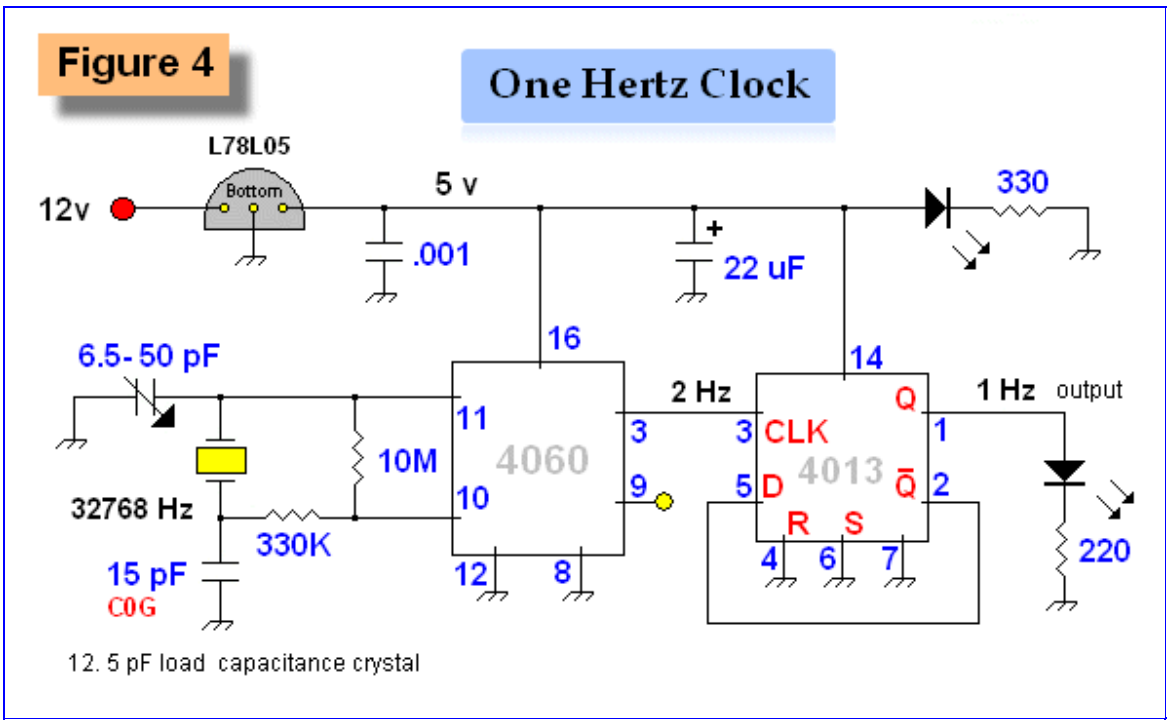
If you are new to digital electronics; (like me) Welcome! Starting small with projects like this one will hopefully lead to increased confidence and problem solving skills for even bigger projects. You can also build the shift registers with J-K flip-flops, but it is more difficult and 4013s or other series D flip-flops are cheap as Борщ (borscht).



Shown above in Figure 3 is how to hook up the XOR gate(s) for 4, 6 and 8 stage pseudo-random number generators. The 6 stage shift-register is of course, Figure 2 above and is presented for reference purposes. The 8 stage version = 1 byte.

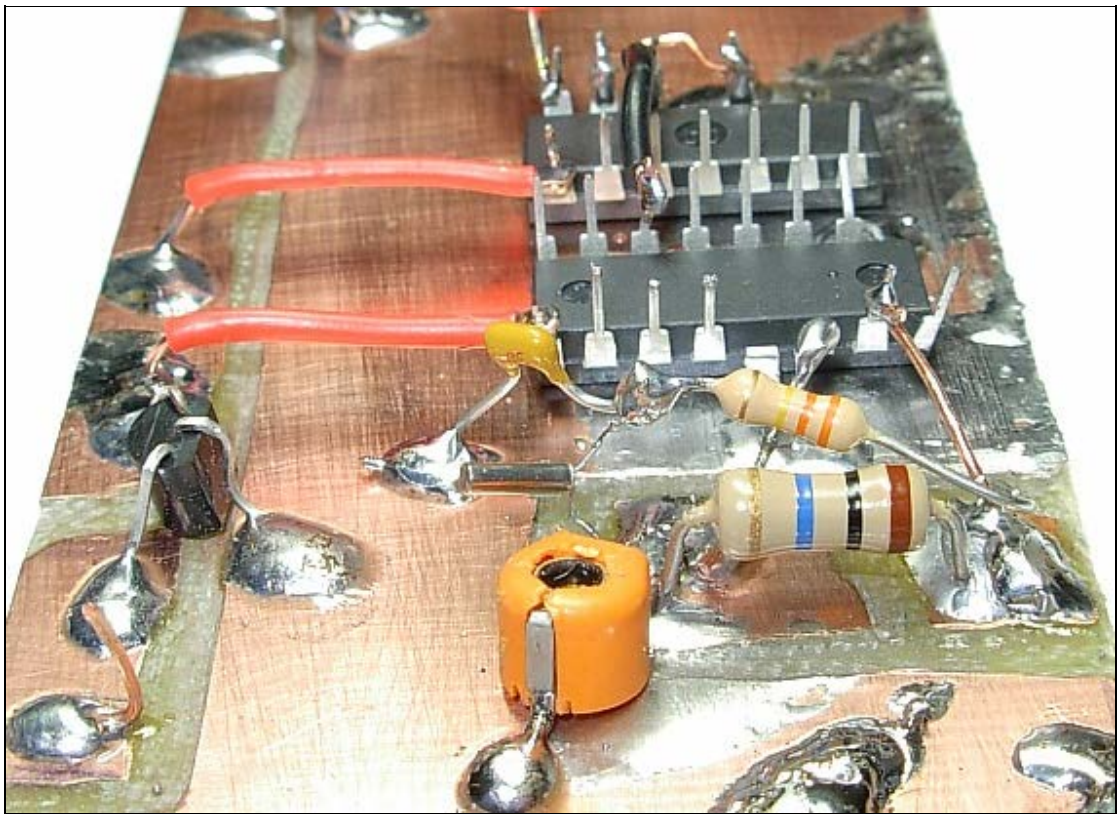
2. One Hertz Precision Time Base

Digital clocks are very interesting. In the past 6 months, 10 -15 RC clocks have been constructed and tested. RC oscillators in the KHz to Hertz range are surprisingly frequency stable. For many projects, a plain RC clock is adequate, however, like in radio design, a crystal controlled time reference is sometimes required. Two examples of projects requiring precision clocks are time of day clocks and frequency counters. Presented is a 1 Hertz clock built from two 4000 series CMOS Logic ICs. Here is a great 4000 series tutorial with pin outs and more.



Shown above in Figure 4 is the complete schematic with an output LED for testing. In the past, the MM5369 17 Stage Oscillator/Divider was popular for hobbyist precision time bases, however, it has gone obsolete. The 4060 ripple counter is a good "modern" replacement, although a different crystal is required. A 32768 Hertz crystal was used and is divided 16384 times to provide a 2 Hz output. The 4060 then drives a 4013 D Flip-flop configured as a divide by 2 to provide a 1 Hz output frequency. Key parts references may be found on the Webmaster's page.

Shown above is the frequency and output waveform when a frequency counter and oscilloscope are (respectively) connected to pin 9 of the 4060 in Figure 4. The 6.5-50 pF trimmer pot is used for calibration. Originally, I used a trimmer cap instead of the fixed 15 pF capacitor shown between the 330K resistor and the crystal. After adjusting this trimmer capacitor for the best looking waveform, I removed the trimmer cap and measured it at 13 pF. I substituted the nearest standard value I had in my parts collection; 15 pF. It was interesting to measure the 4013 output frequency at 1 Hz.



A close up photo of the 4060 oscillator/divider breadboard. The 10M resistor used was a 1/2 watt rated R as I have dozens of these in my parts collection. You can see the tiny cylindrical crystal just above and left of the orange Murata trimmer capacitor. It is oriented horizontally. This is a useful time base for the QRP workshop.

3. 10000 and 5000 Hz Multivibrator Clock

It is fun to occasionally build circuits using discrete semiconductors rather than with ICs. A 5000 Hz digital clock was needed for an experiment. It

was decided to use multivibrators for the basic oscillator and a divide by 2.

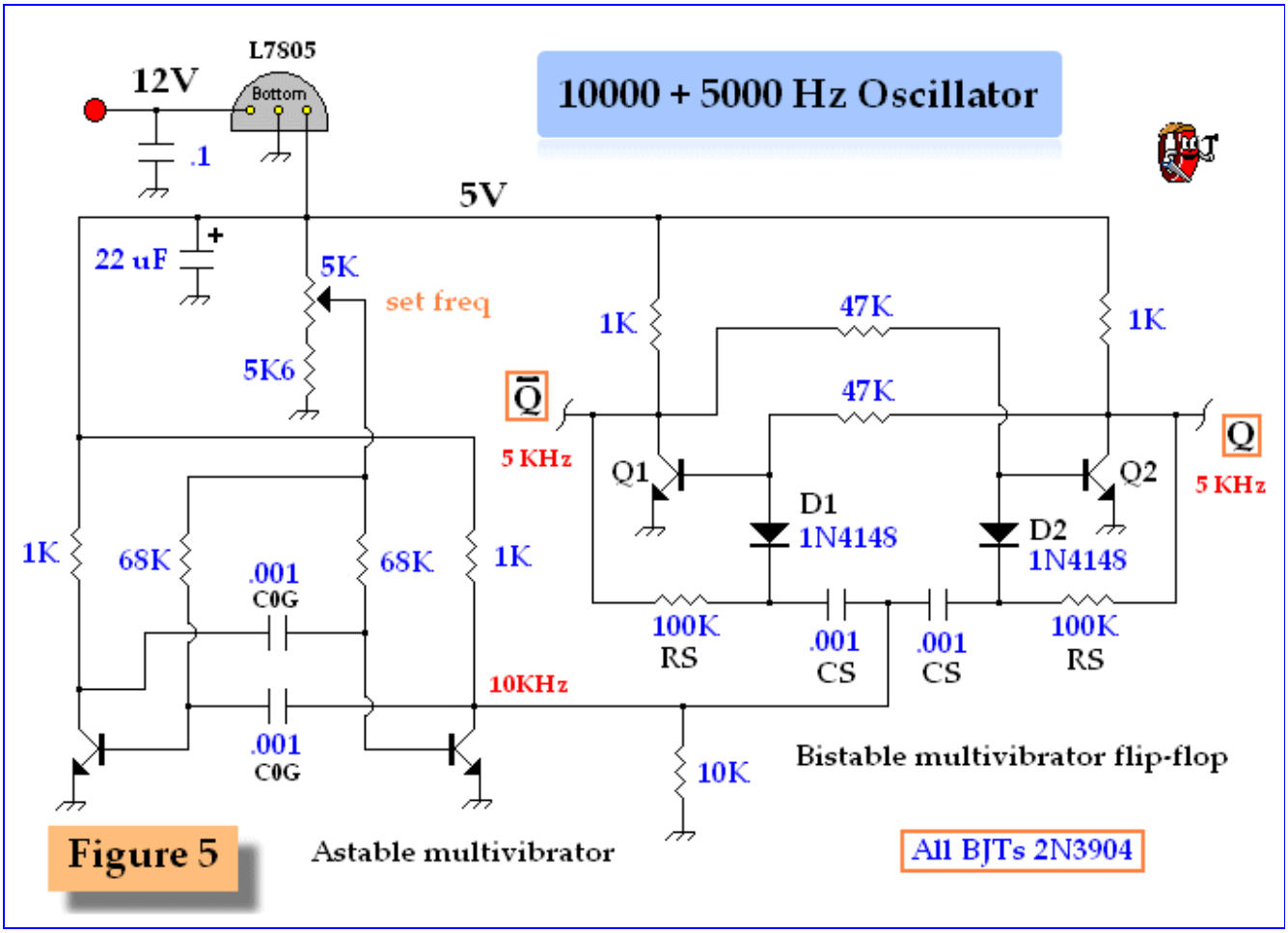
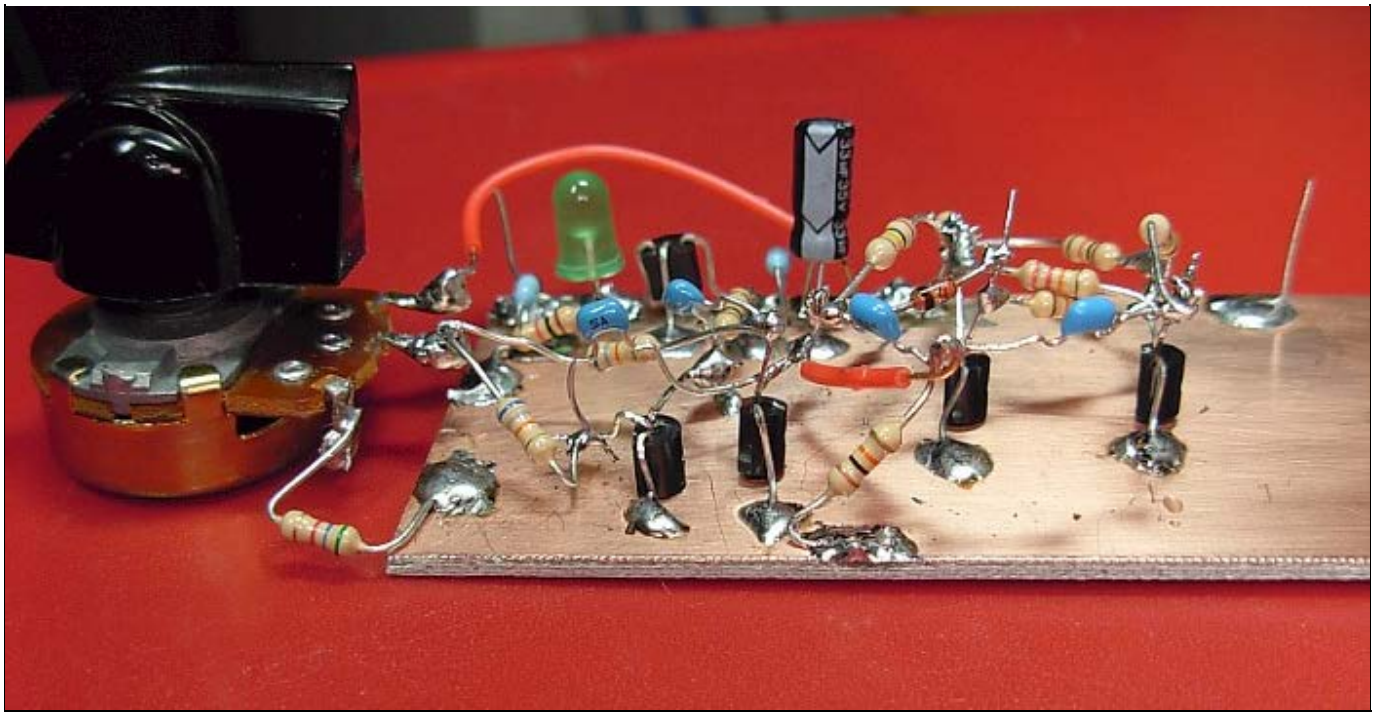


Figure 5 is the entire circuit. The tuning range of the astable multivibrator was about 7060-10650 Hz. The 5K pot was slowly adjusted until 10000 Hz was measured in a frequency counter. Following testing of the astable multivibrator, the flip flop was built and examined. Astable multivibrator function has been discussed previously on this web site.

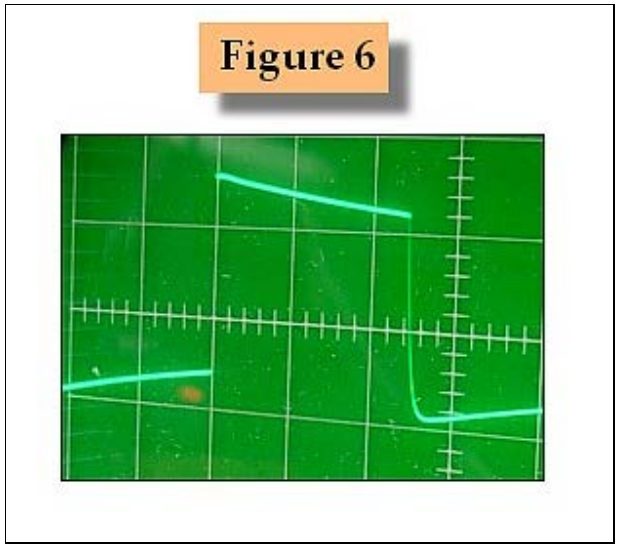
Please refer to the bistable multivibrator. It is a one input circuit set up for toggle or flip-flop operation. Negative edge pulses applied between the two 0.001 capacitors will cause the binary state of Q1 and Q2 to change to the opposite state. The multivibrator circuit is made up of Q1, Q2 and the 47K and 1K base and collector resistors respectively. The other components D1, D2, the RS resistors and CS capacitors comprise a steering circuit to generate the proper response to the negative edge pulses. When a negative input pulse arrives, it is guided to the base terminal of the ON transistor, but prevented from reaching the base terminal of the OFF transistor.

In order to study this circuit at DC, I temporarily exchanged the 0.001 timing capacitors in the astable multivibrator with some 22 uF electrolytic caps to slow it down. Referring back to the bistable multivibrator, let us assume that Q1 is OFF and Q2 is ON. The collector voltage of Q1 is high (cut off). The collector voltage of Q2 is low (saturation). The Q1 collector is connected to the cathode of D1 by the 100K RS resistor. The cathode of D1 is reverse biased by the high Q1 collector voltage and also because its anode is held close to 0 volts by the 47K resistor connected to the collector terminal of Q2. It would take a very strong negative input pulse to forward bias D1 enough to reach the Q1 base terminal. The Q2 collector voltage is nearly 0 volts and therefore the D2 cathode has little to no reverse bias voltage via its RS. Thus, any small amplitude negative input pulse will cause D2 to become forward biased, reach the base of Q2 and drive Q2 OFF. Once Q2 switches off, in turn Q1 is toggled ON and its collector voltage goes low. The large reverse bias on D1 disappears. However, Q2 is now OFF and D2 will now be strongly reverse biased which will steer the next negative input pulse to the base of Q1. This is the basis of the circuit's negative edge flip-flop operation.

In another experiment, I changed the .001 C0G capacitors of the astable multivibrator to 470 pF. This gave a usable range of 22968 to 14832 Hertz (11484-7416 Hz at the Q1 and Q2 output) . Looking at the output of the flip-flop in the oscilloscope; at the higher frequency range, the flip-flop could not keep up and failed to divide by 2. I found experimentally that the time constant of each of the CS and RS components seemed to be the problem. When the CS capacitors were also decreased to 470 pF, the flip-flop worked properly.
As you increase the flip-flop operation frequency, speed up bypass capacitors might also be required across the 47K base resistors of Q1 and Q2 . A suggested starting value to try is 220 pF. Some builders also bypass the resistors in the RS steering circuit at higher frequencies, however, this is getting a little crazy. It is really important to look at the output waveform in the oscilloscope to ensure reasonable performance.



Shown above is the Figure 5 breadboard prototype.



5 KHz output waveform of Q2

4. One KHz Digital and Analog Oscillator

A 1 KHz oscillator with 5 volt digital outputs 180 degrees apart and an analog output was sought. The frequency had to be near to, but not exactly 1000 Hertz. A major question to answer was how much low pass filtering is needed to remove the odd harmonics from digital circuits?

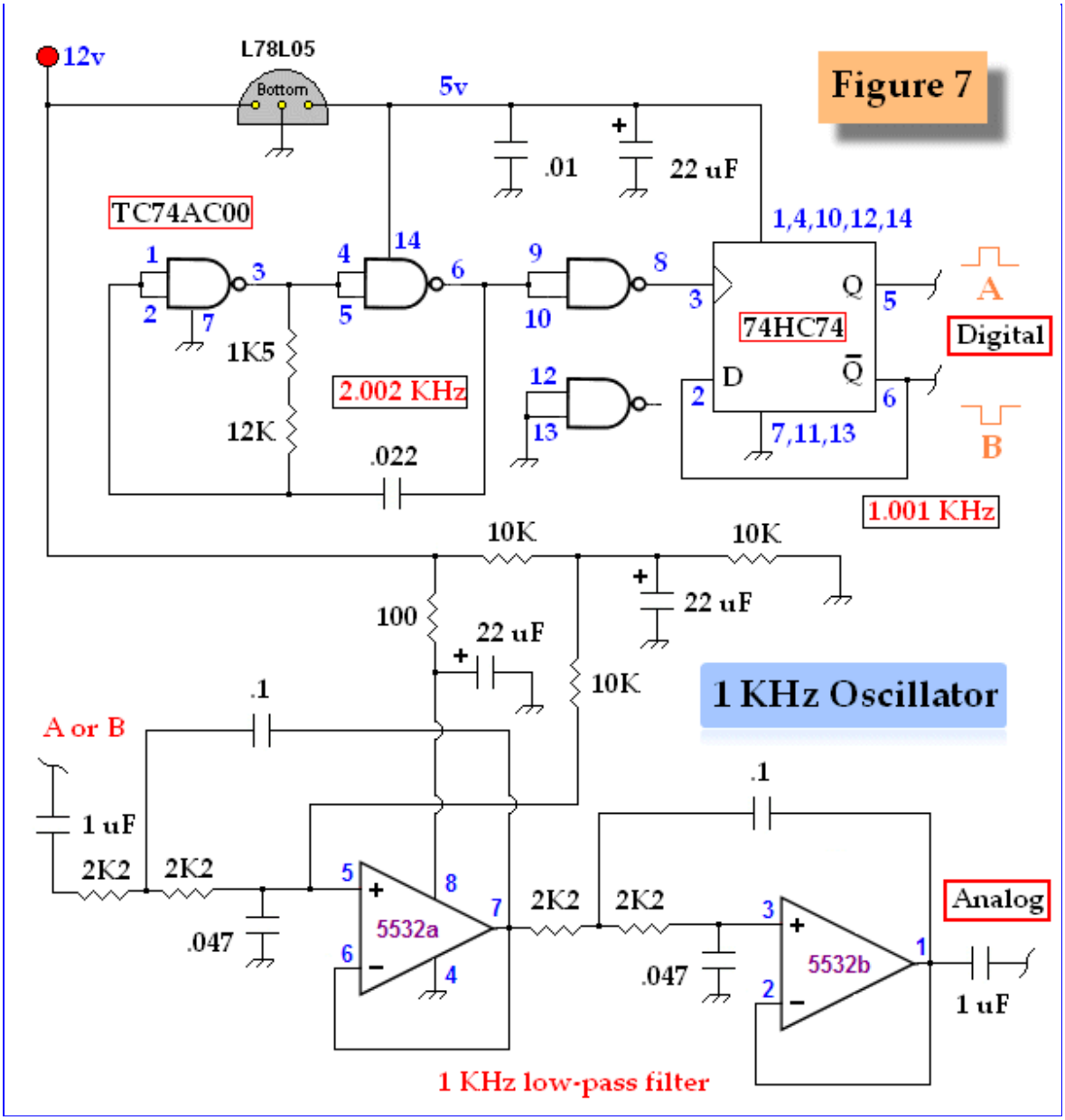
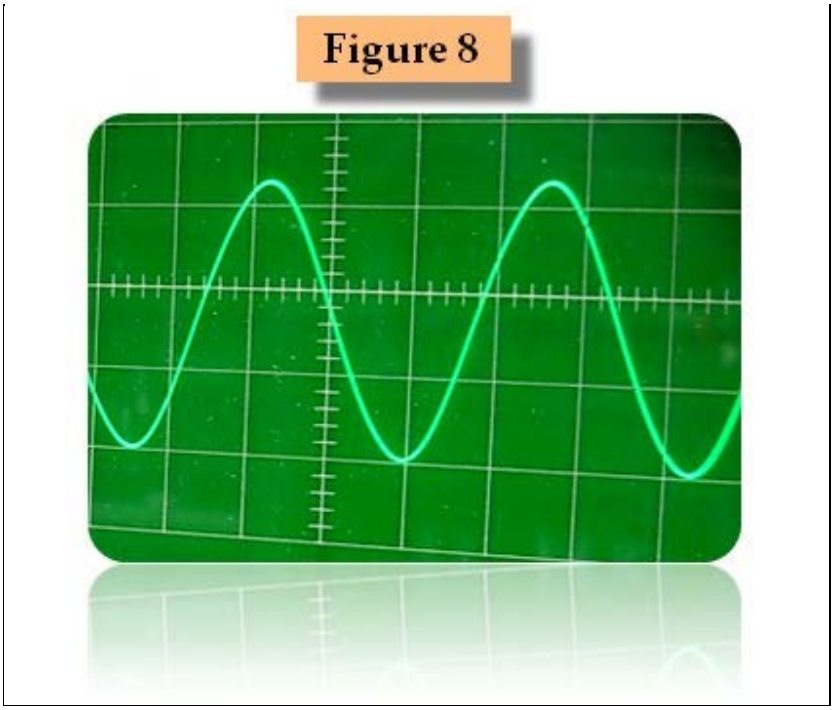


Figure 7 shows the complete schematic. NAND gates from a 74AC00 were wired as inverters and with the 13.5 K resistance and a 0.022 uF polyester capacitor, the frequency was 2002 Hertz. To improve the digital waveform and get the desired 2 outputs, a D flip-flop was used. The output frequency was 1001 Hertz. The digital part was completed!

For the analog filtering, active low-pass filters were tried, and in total 4 poles with a 1 KHz cuff off worked reasonably well. The filter uses the 5532 op-amp with common vales capacitors and resistors. Poly"something" caps were utilized.



In Figure 8 is the output waveform of the low pass filter stages. A pretty nice sine wave was achieved and this oscillator could see duty for testing audio amplifiers. The scope was photographed at an angle to avoid the camera reflection and this distorts the sine waves a little.

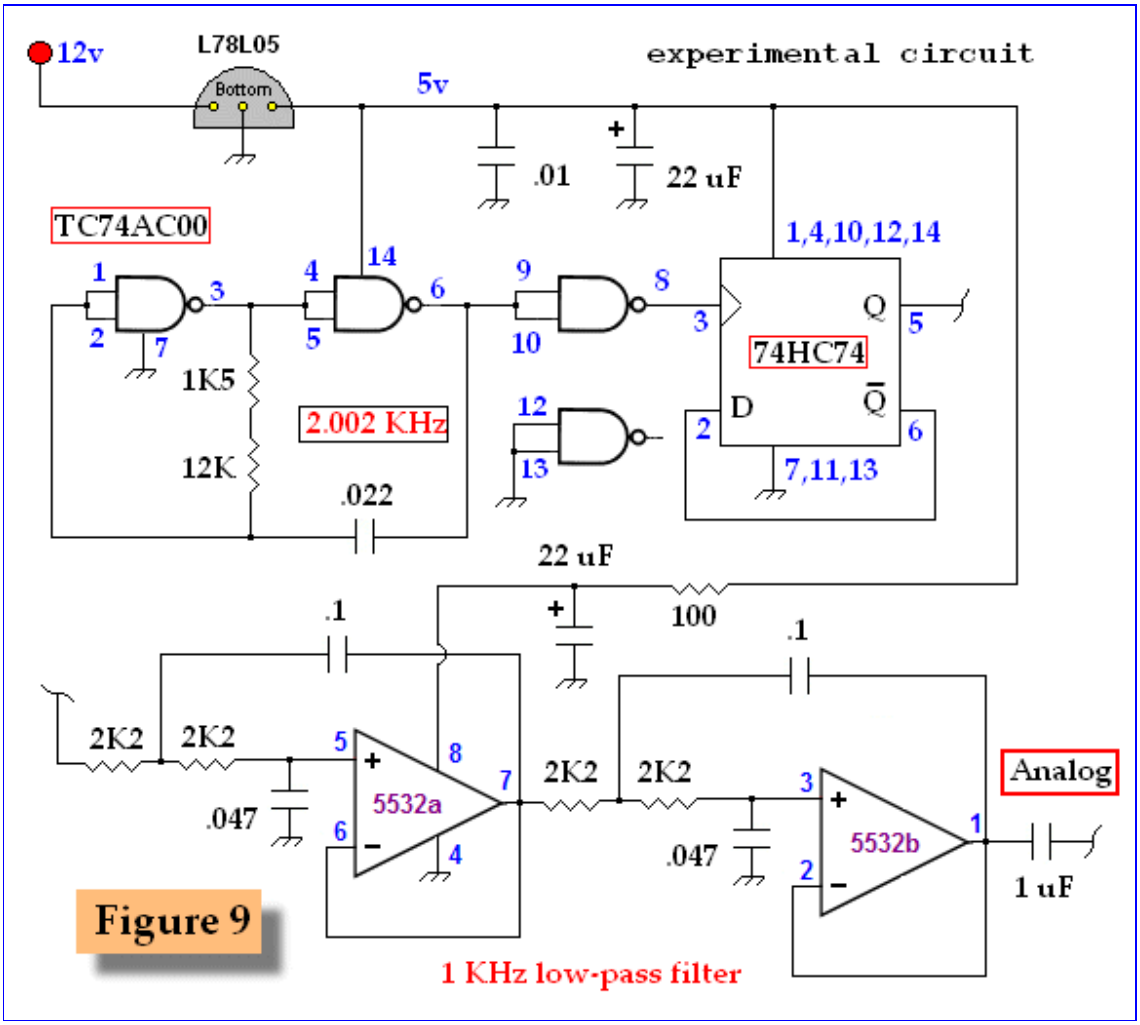


Figure 9 depicts an experiment with the op-amp biasing. If the op-amp is run at 5 volts VCC, the bias requirement is 1/2 VCC or 2.5 volts. The DC voltage at the output of the D flip-flop was 2.55 volts. The 2K2 resistor was connected directly to this output and this eliminated the VCC/2 resistor bias network and a coupling capacitor.

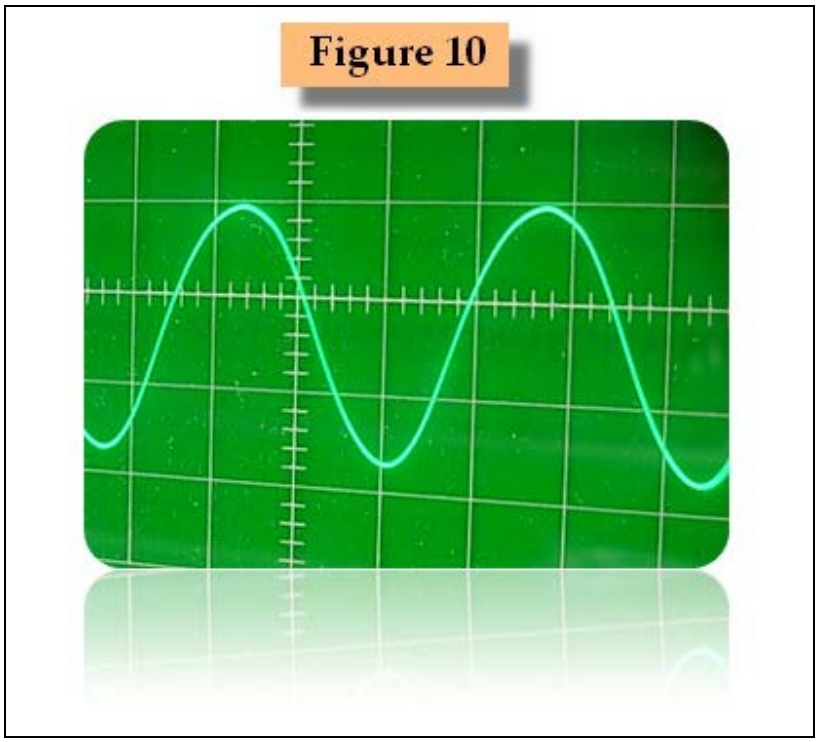
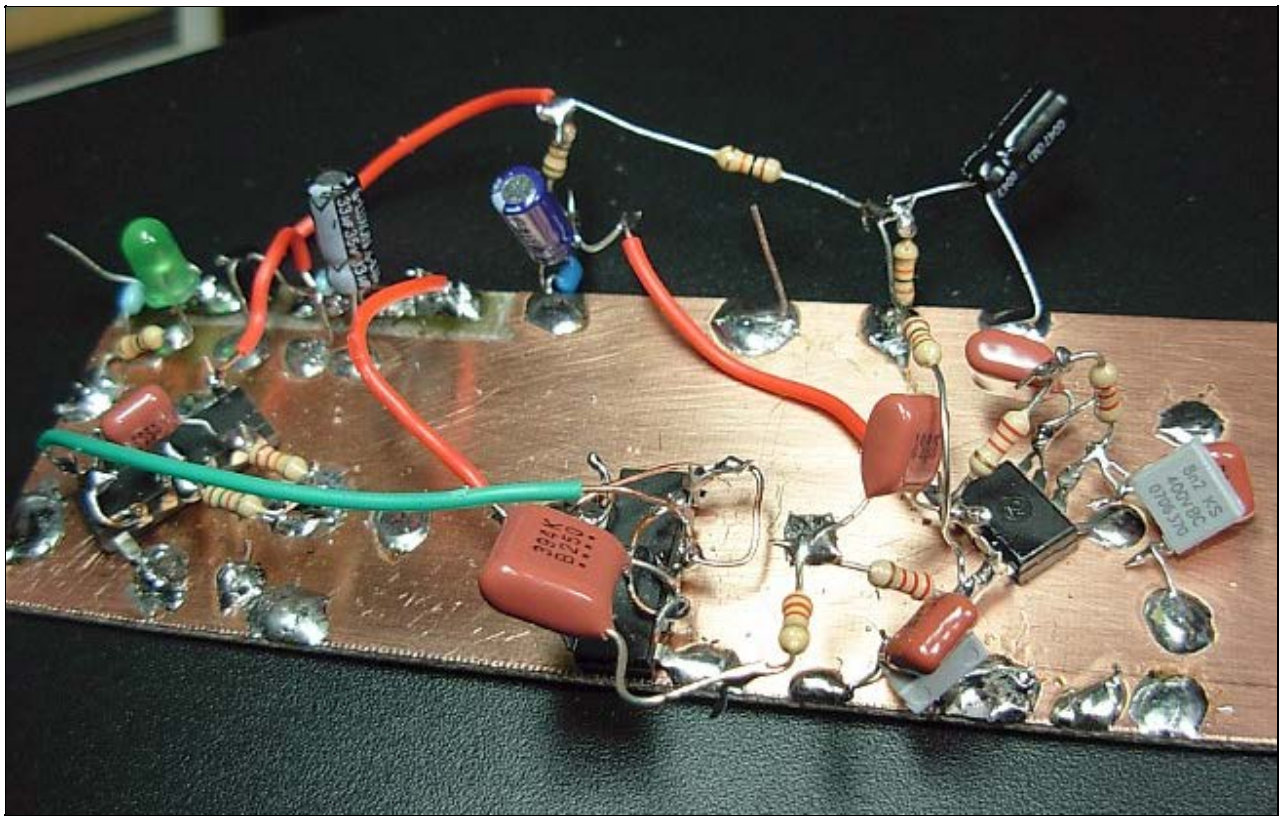


Figure 10 shows the output waveform of Figure 9. The AC waveform has harmonic distortion and thus the Figure 9 circuit will not be kept nor utilized.



The Figure 7 breadboard. A 0.39 coupling capacitor (not 1 uF) was used between the D flip-flop and 5532a in this particular version. Unfortunately, no 0.047 uF caps were available for the low-pass filters and therefore a 0.039 plus a 0.0082 were placed in parallel for each of the .047 uF caps.

5. One KHz Low Distortion Signal Generator

Although I own a variable frequency wein bridge oscillator, it has been been set to 1 KHz for 2-3 years and is large and temperamental. It was decided to make a low distortion sine wave oscillator for just this one frequency. The circuit will be placed in a box along with another signal generator.

There are a number of ways to build signal generators using op amps. Countless example circuits may be found on the World Wide Web and some of them are really fantastic. Chose whichever method works best for you. Some might find my circuit to be overkill, but to each his own.

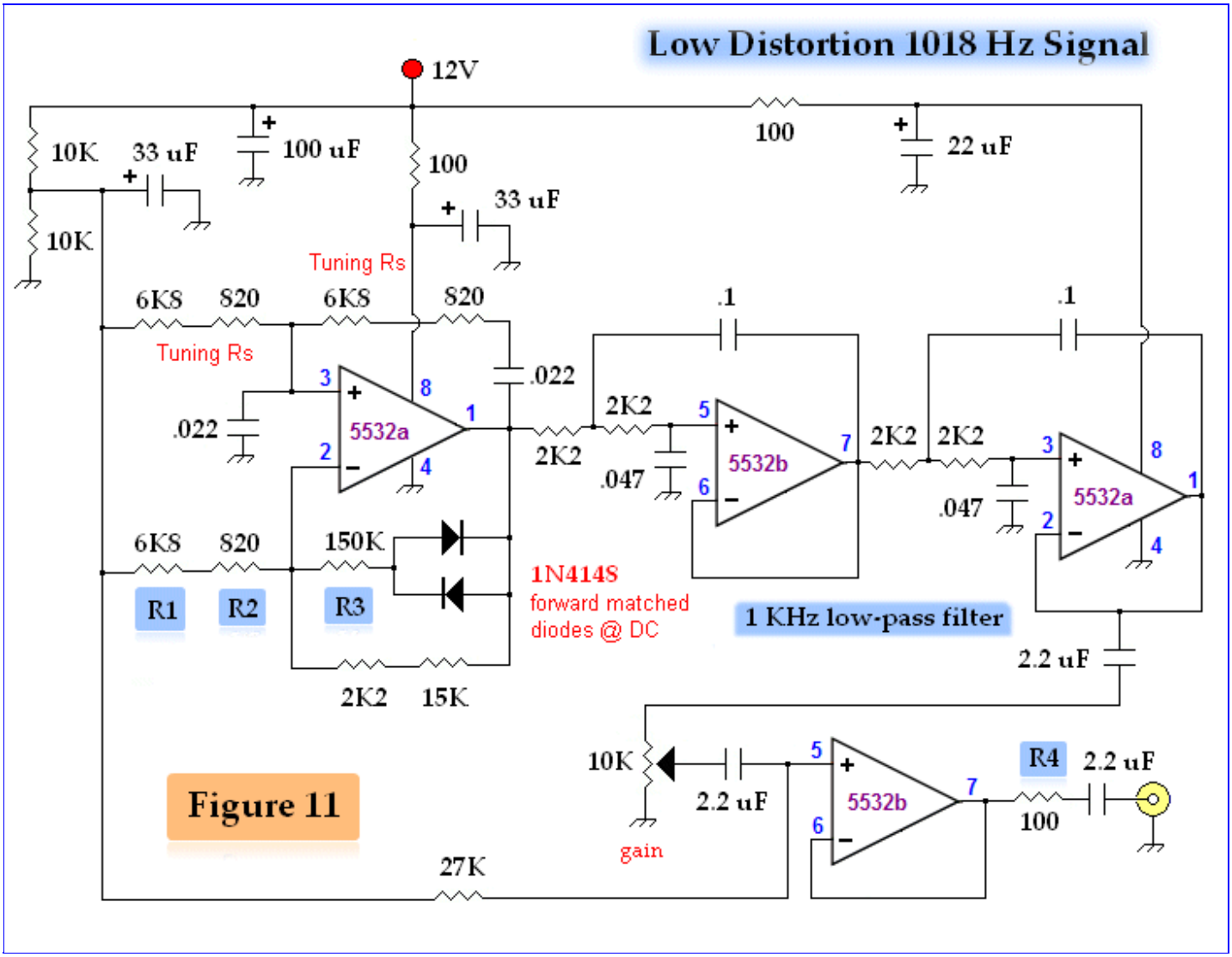


Figure 11 shows the entire circuit. The Wein bridge oscillator is from EMRFD and was designed by Wes, W7ZOI.

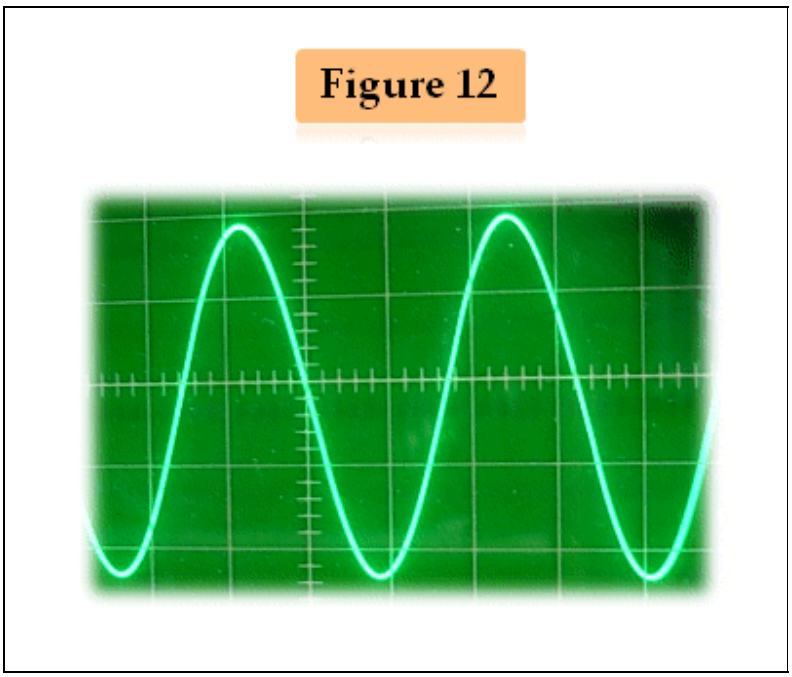
A 0.22 uF capacitor was chosen for the tuned circuits. Using resistors from my parts collection, 1018 Hz was the closest I could get to 1 KHz. This is the 6K8 + 820 ohms resistors labeled "tuning Rs". Other values were tried. For example, a single 6K8 gave 1142 Hz and a single 8K2 gave 939 Hz.

When the circuit was first built, I used a 10K on the VCC/2 bias point to pin 2 and a 22K feedback R from pin 1 back to pin 2. The sine wave had mild distortion. By experimentation, it was learned that the resistance from the VCC/2 bias point to pin 2 significantly affected the waveform purity. The 2K2+15K plus the R1 + R2 resistance values shown were determined by using a potentiometers rather than fixed resistors. Care was taken to adjust the feedback resistance from pin 1 back to pin 2 to keep away any overdrive distortion. I do not understand this, but even changing the 820 ohm R2 to 570 ohms, altered the sine wave purity.

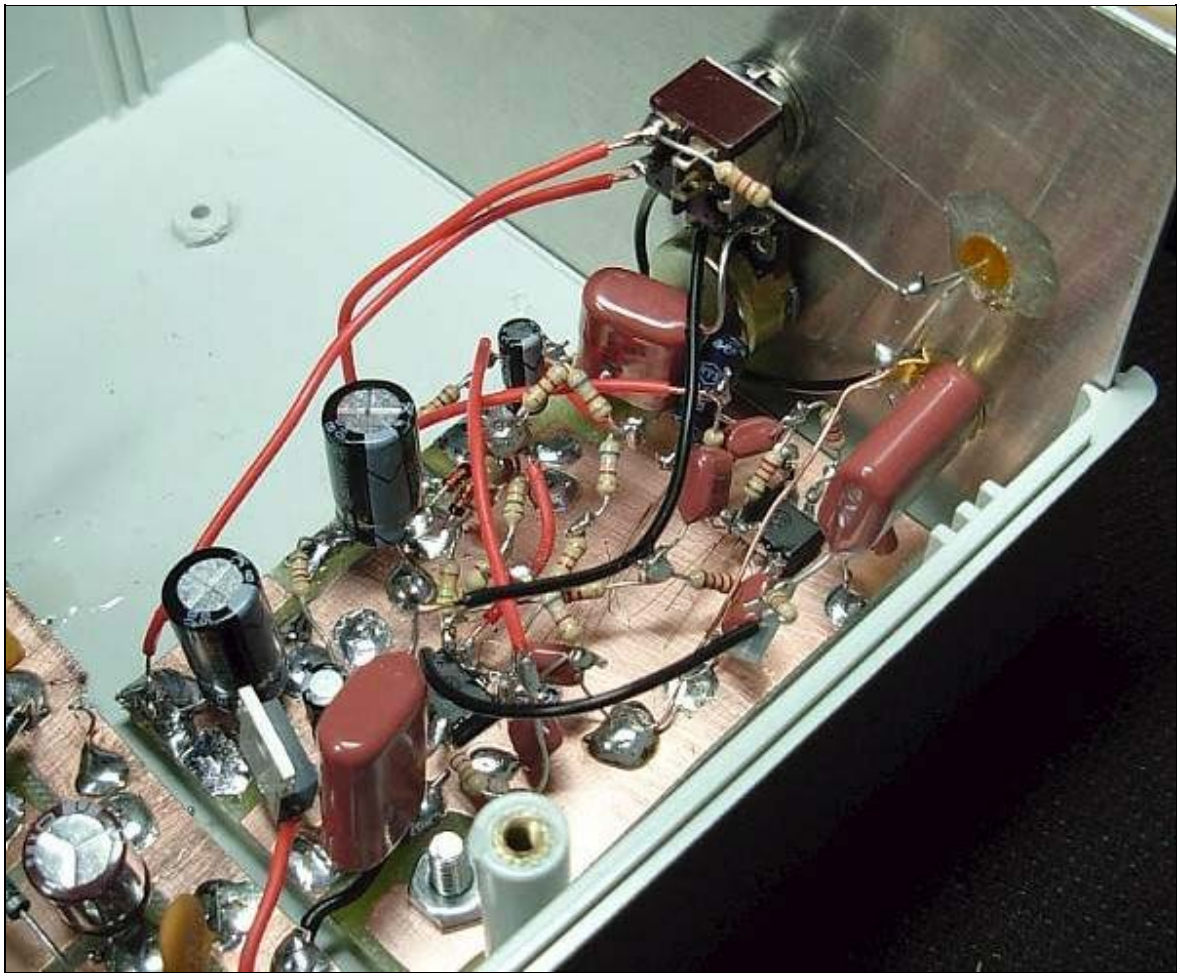
The best looking sine wave came when the resistance from the VCC/2 bias point to pin 2 was the same as the tuning resistance; 6K8 plus 820 ohms. Later, the pin 1 to pin 2 feedback resistance was chosen for an unclipped; waveform with a reasonable output voltage using a potentiometer. The potentiometer was removed and measured at 17.1K, thus the 15K + 2K2 were soldered in. It was also discovered that by increasing R3 from 56K or 100K to 150K slightly improved the waveform.

The Figure 7 low-pass filter was connected to the main oscillator as shown. The final op-amp stage was used as a buffer between the low-pass filter and the gain control. R4 is used to set whatever output impedance you choose. Practically speaking, it could be any value between 47 and 620 ohms. Many AF oscillators have an output impedance of 600 ohms and 620 is the nearest E24 standard value. For my project, a 100 ohm

R4 was chosen. Output peak to peak voltage is 0.0 to 4.84v continuously.

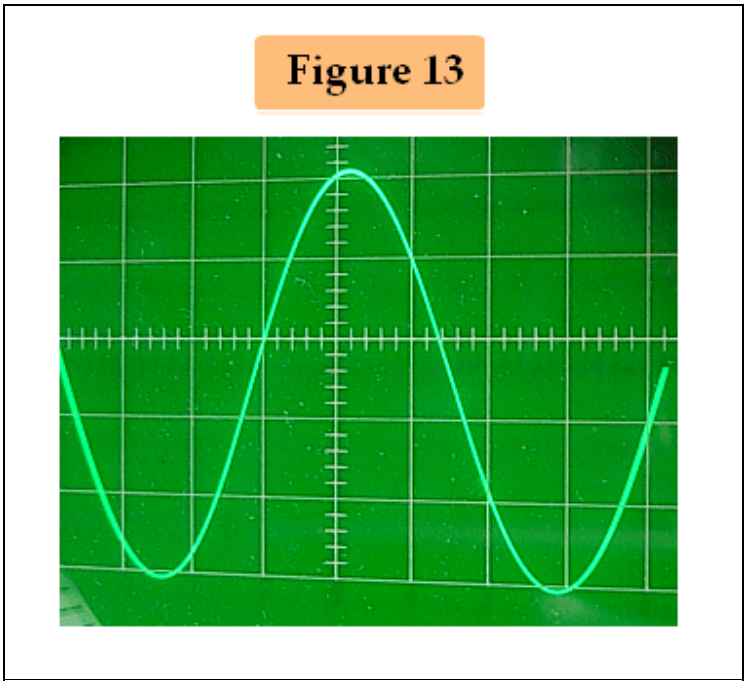


Here is the raw output of the basic Wien bridge oscillator. It is hard to photograph well, but it is stellar to say the least.



The Figure 11 breadboard mounted in a plastic Hammond chassis. The voltage regulator seen to the left of the bottom polyester cap is a 7812. This project has its own regulated power supply. Other view of Fig 11. Two 10 Megohm standoff resistors were used to help support all the resistors soldered to pins 1, 2 and 3 of the main oscillator.

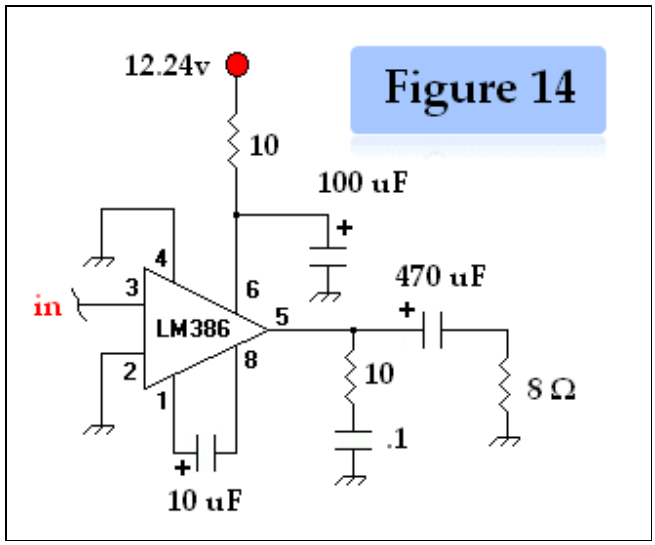
A front view photo of the AF oscillator. A separate 7 plus 14 MHz oscillator circuit and controls will be placed on the right hand side of this box. The orange power ON indicator LED was epoxy glued into the chassis hole. Putting circuits in cabinets is one of the most expensive aspects of homebrew construction. One must be ever vigilant for bargain chassis boxes and hardware to keep costs down. Techniques such as gluing in the LED rather than purchasing a separate holder and recycling knobs and switches are also practiced for cost containment.



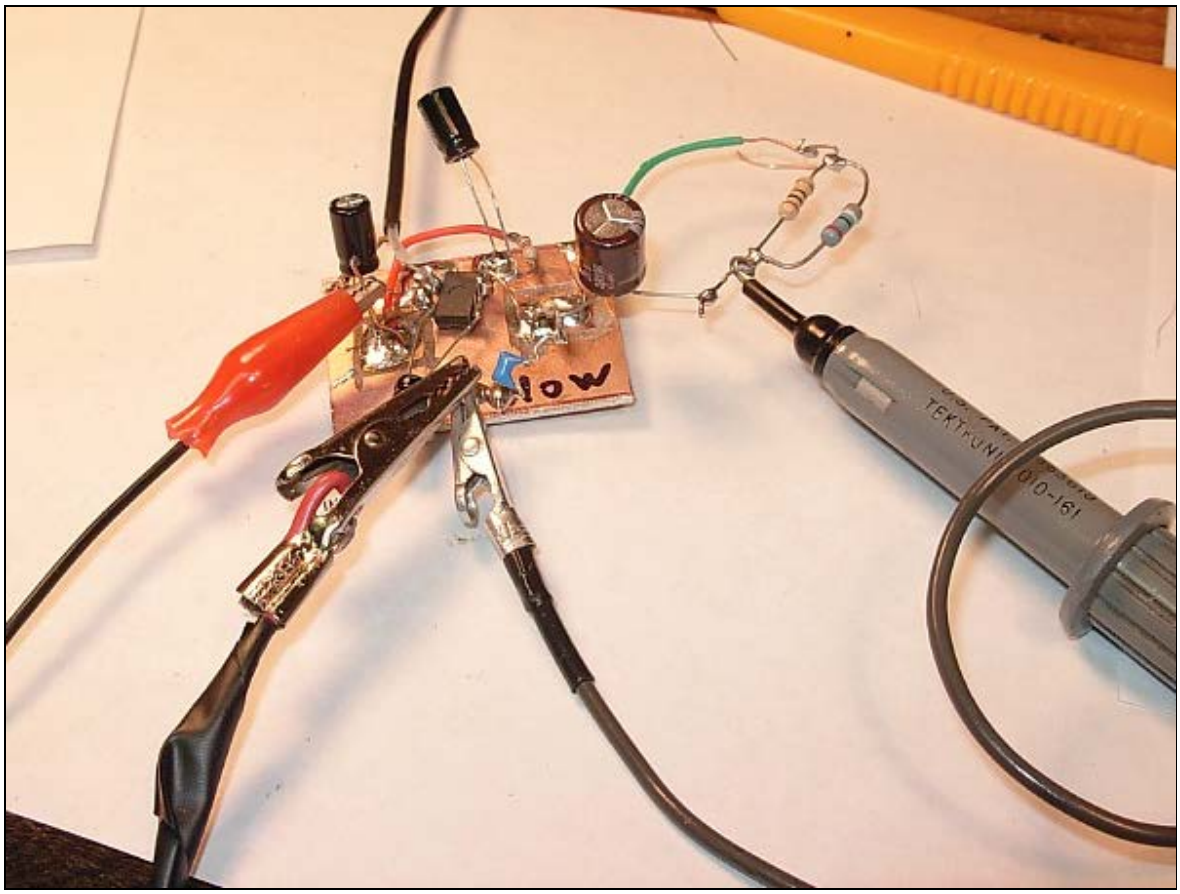
The output of Figure 11 is shown in Figure 13. This is the best sine wave seen ever on my scope. I looked at it closely and there is no change in line thickness or symmetry anywhere. It inverts with no change on the scope. Testing audio amplifiers will now be much more fun.

6. LM386 Power Experiments

The LM386 is an IC audio amp that has been used in thousands of hobbyist projects over the past 2 decades. By adding a capacitor +/- a resistor between pins 1 and 8, this device's internal gain can be changed from x20 to up to x200.



Test circuit schematic in Figure 14.



The experiment breadboard is shown above. A very standard configuration. The amplifier drove an 8 ohm, 1 watt resistive load.

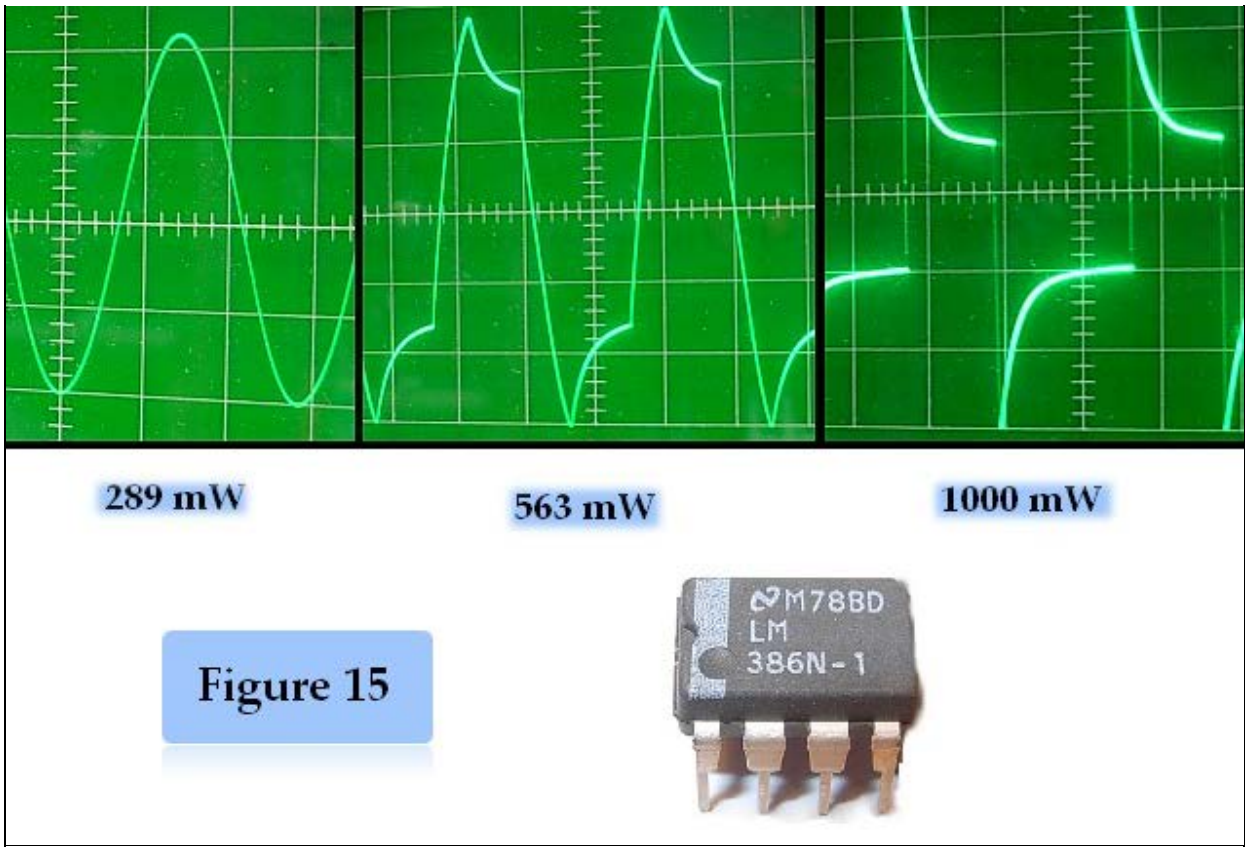


Figure 15

Over the years, I have noticed some kit sellers and project authors claiming that their LM386 based AF stages gave 1 or occasionally even 2 watts of output power into an 8 ohm speaker. This was confirmed on the bench. This device will output 1 watt into 8 ohms at 1018 Hertz with little problem. However, this is clearly 1 watt of square wave distortion.

The quiescent current of the LM386 was around 7 mA. The signal generator gain was increased until the first signs of distortion appeared. The gain was then backed off a little so a pure sine wave was observed in the oscilloscope. The current was ~ 155 mA and the measured power was 289 mW. Please refer to Figure 15 for the 289 mW sine wave. This was the clean signal power of the LM386 on my bench. The output waveforms at 563 mW and 1 watt are also shown. Extreme harmonic distortion occurred above 300 mW. This device will draw 240 mA or more when driven and clipping hard. It is not my intention to malign the LM386. It is a useful part, albeit a little dated. Its AF gain capability versus size is something to behold. Many builders have moved to the TDA7052 audio amplifier IC, or like myself, build their own low noise audio power amps.

The 12.24 volt DC supply and the 1018 Hz AF audio oscillator used in these experiments.

Time Domain Output from a Diode Ring Mixer

22 Dec 2009, w7zoi

Figure 1. Some available experimental modules.

Some folks wonder about the output that they should see on their oscilloscope when looking at the output from a diode ring mixer. There is no set, pat answer. The output
can change dramatically as levels, frequencies, and even terminations are changed. This complication is illustrated here with a few screen shots, taken with a Rigol DS1052E 50 MHz bandwidth digital storage oscilloscope. The experiments started with the following pile of modules. Your collection will probably differ.

Figure 2. The inside of the module containing a Mini-Circuits SBL-1 diode ring mixer. This is a standard part that is essentially generic.

The first experiment was to set up a pair of 10 MHz signal sources. One was from a homebrew generator, shown below.

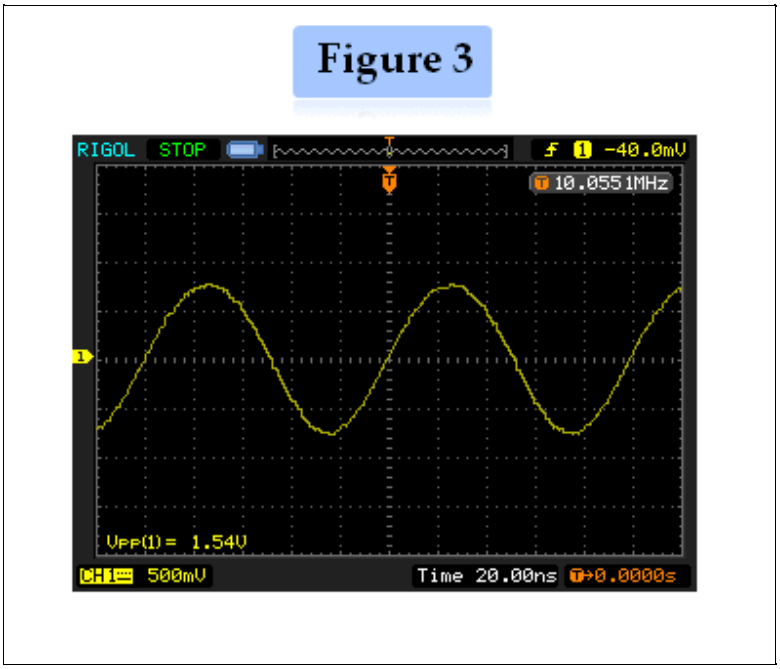


Figure 3. The 10 MHz signal from a homebrew signal source. The 1.54 volt peak to peak signal is applied to a 50 Ohm terminator at the oscilloscope. The delivered power is then +7.7 dBm.

This signal was filtered with a 14 MHz low pass circuit. This caused the amplitude to drop by 0.2 dB. The source was then attached to the LO (local oscillator) port of the SBL-1 mixer.

Figure 4

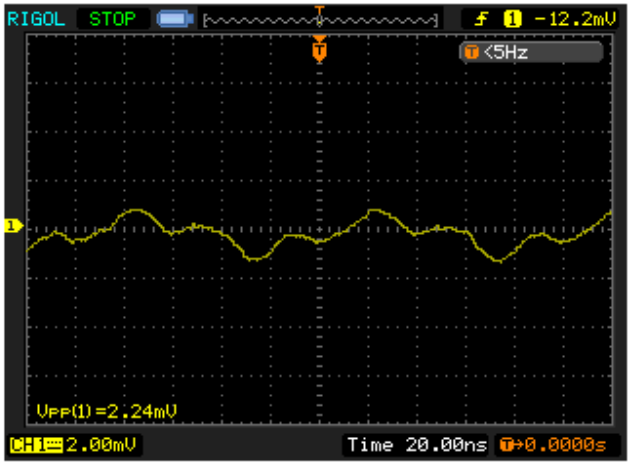


Figure 4. The IF output from the mixer when there is nothing attached to the RF port. Note the scope sensitivity of 2 mV/div.

Next, we attached a 50 Ohm terminator to the R mixer port.

Figure 5

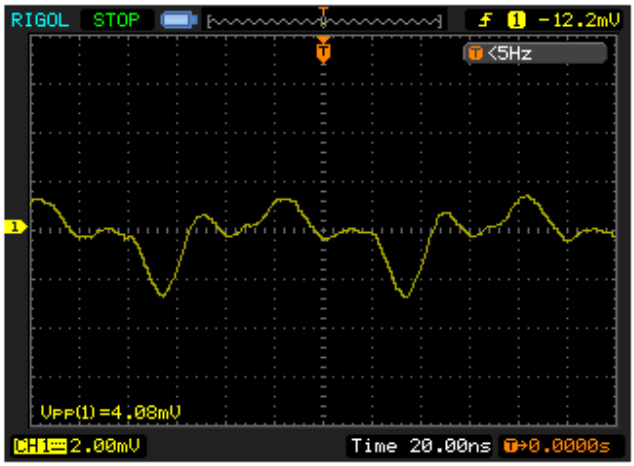


Figure 5. The IF output with LO drive, but without an R signal. But the R port is now terminated. This waveform, when compared with Fig 4, shows just how sensitive the mixer can be to termination.

In the next experiment, a -20 dBm signal was applied to the R port. The frequency was very close to the 10 MHz LO that is still present.

Figure 6

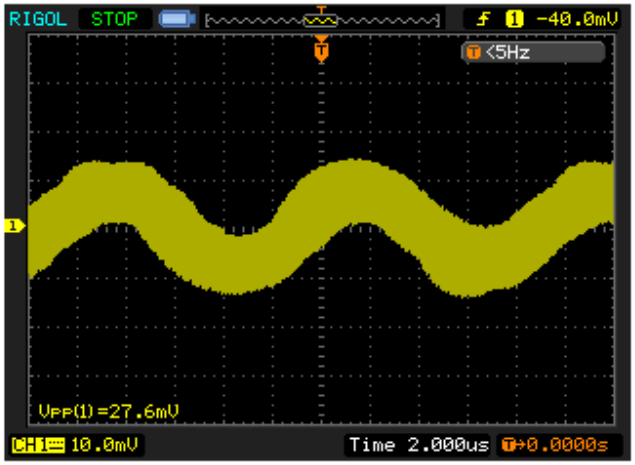


Fig 6. There are two dominant signals from the mixer. One is a low frequency at 100 kHz. But this is accompanied by a high frequency of about 20 MHz. These two outputs, a sum and difference frequency, are expected from any mixer.

A filter can isolate the two dominant outputs. This is shown below where a 500 kHz low pass filter is inserted in the line between the mixer and the oscilloscope. A 6 dB pad is between the mixer and the filter, for direct insertion would upset the termination of the mixer.

Figure 7

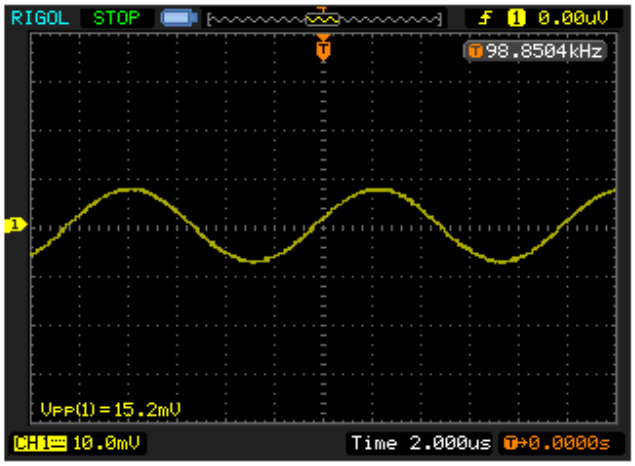


Figure 7. The output of about 100 kHz after a low pass filter is inserted in the mixer output.

The next experiments emulate a SSB transmitter. We start with a signal at 11.06 MHz with strength -20 dBm. (This is a common IF used in homebrew SSB transceivers such as the BITX-20.) This is applied to the mixer R port. The L port is driven with a +7 dBm signal at 3.19 MHz. The LO signal is low pass filtered to attenuate harmonics, a measure that is probably not necessary, but the filters were there. The IF output is shown below.

Figure 8

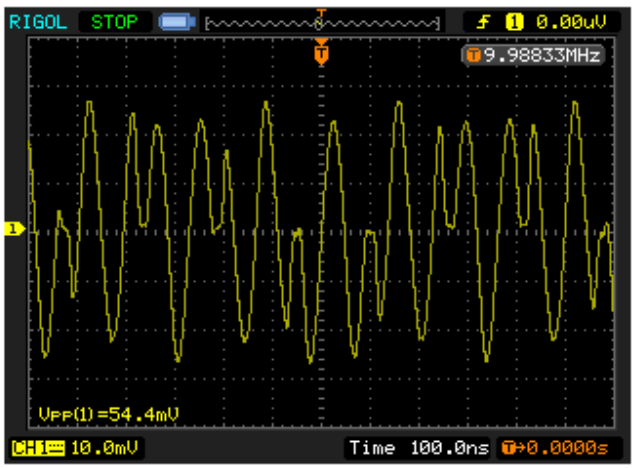


Figure 8. Time domain output of a SBL-1 set up as the “transmit” mixer in a SSB rig. This is far from the “perfect sine wave” that some folks tell us we should observe. This waveform contains many different frequency components. The counter output should not be interpreted to have any meaning. (I should have turned it off.)

The signal of Fig 8 can also be viewed with a spectrum analyzer. This is shown below. This measurement was taken with the August 1998 QST Spectrum Analyzer and not the FFT routine in the DSO. The Rigol scope has a nice display for an analyzer.

Figure 9

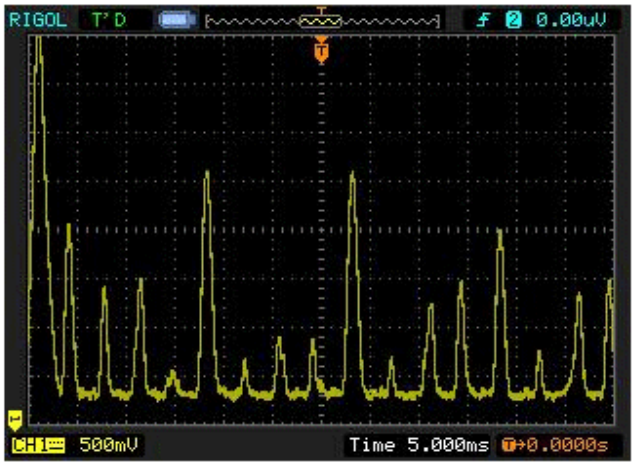


Figure 9. Spectrum of the signal shown in Fig 8. The largest signal on screen is that at the left, which is the spectrum analyzer zero spur. This is a spurious output that is typical of most SA systems. The desired signal at 11.06+3.19=14.25 MHz is just to the right of center. But the image is also present at the different frequency of 7.87 MHz at about 3 major divisions from the left edge.

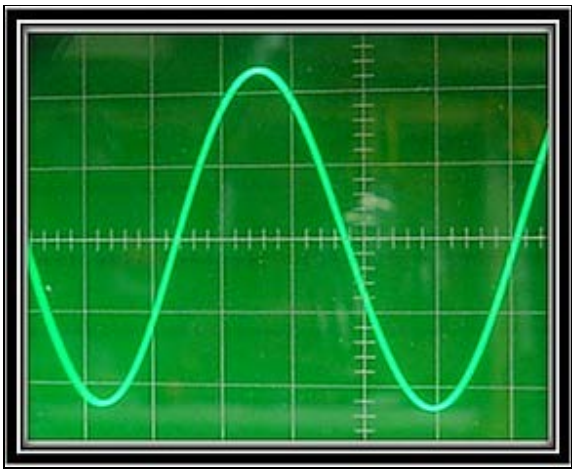
Alas, I didn’t find a 14 MHz bandpass filter in the junk box. Such a filter would have allowed selection of the dominant 14 MHz component while attenuating all the rest of the junk shown. The many other signals are the result of harmonic mixing. That is, we observe IF outputs at N x FLO +/- M x FRF where N and M are integers. Some of these spurious outputs can be quite strong with diode ring mixers. They are best avoided with high frequency LO signals. In this case, a LO at 14.25+11.06=25.31 MHz would produce a much cleaner output spectrum. It is much easier to obtain LO stability with an oscillator built at 3.19 MHz.

Bottom Line

It is not reasonable to have a well defined, predictable time domain (i.e., normal oscilloscope) output from a mixer. The exact results depend upon too many variables. A spectrum analyzer can be used to garner much more information.

Low Noise Crystal Oscillators

Introduction

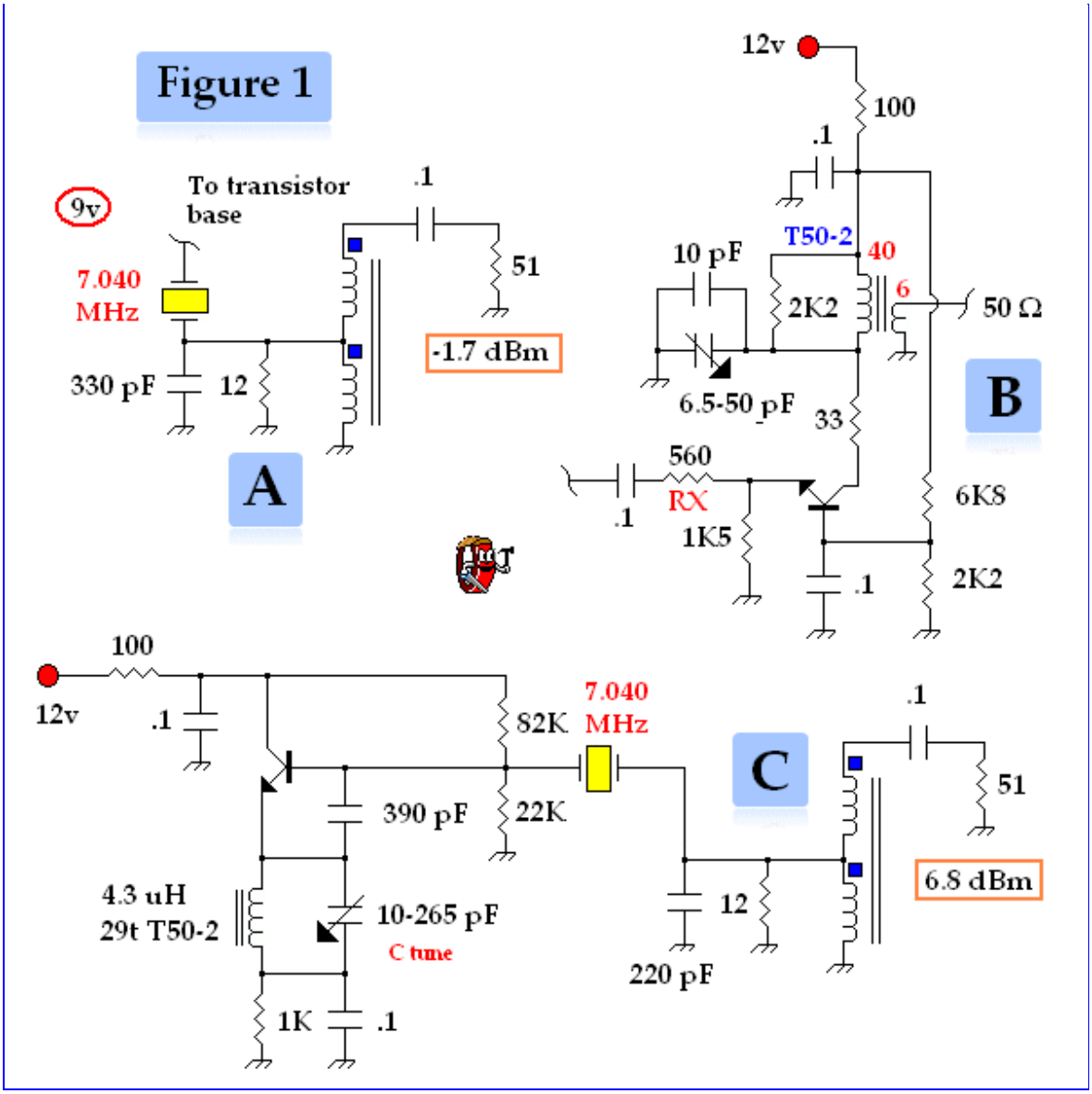


Some experiments were conducted to build a low noise crystal oscillator with 50 ohm output at 7 and 14 MHz for the test bench.

Some readers might wonder why build such an oscillator? Although a variable RF signal generator is an important bench tool, it is also nice to have a fixed frequency signal source on your favorite HAM band. This RF source can be connected to other 50 ohm modules such as band-pass filters, diode ring mixers or feedback amplifiers to conduct experiments at a whim. The project goal was a low noise oscillator with 2 outputs so it could drive a divide by 2 flip-flop for 40 meter band digital mixer work, or fundamental frequency use on the 20 and 40 meter bands.

Presented are some experiments carried out to realize this goal. Only some of the better experiments and circuits are shown. An additional circuit was added Jan 31, 2010.

First Steps



The 7.040 MHz crystal used is an AT cut, HC6-U holder part made locally by West Crystal. I reviewed the information concerning crystal oscillators in Chapter 4.5 of EMRFD and then started melting solder. The oscillator output was extracted as described in EMRFD Figure 4.24. The output is low distortion, low impedance and low gain. Like in most experiments, I built it, measured the DC voltages and then looked at the AC voltages in the oscilloscope.

In order to measure the output, a 51 ohm load was transformer coupled as shown in Figure 1a. I am uncertain if this was a good method for power measurement, however, it allowed comparison of the experimental circuits. As shown, the output with a 9 volt regulated power supply is low; -1.7 dBm. Still, a circuit like EMRFD Figure 4.24 can be used with a variety of crystal frequencies and has great utility.

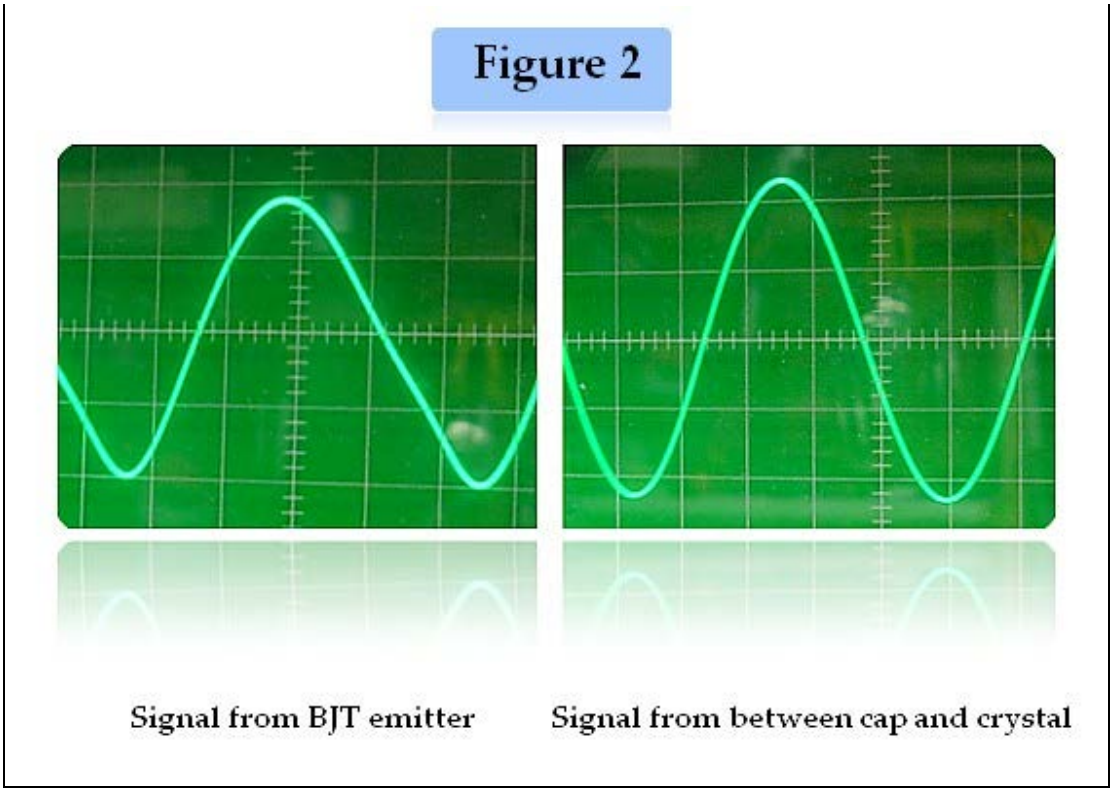


Figure 2 shows two output waveforms taken from my crystal oscillator. When a signal is taken from the emitter of the main oscillator transistor (what we typically do) harmonic distortion occurs as shown in the above left. Actually, the distorted waveform photo above left looks better than most do. Typically, they look like this. Many builders will just place a low pass filter on such an oscillator's output and be very satisfied with the harmonic content in their signal. Certainly this is a good, common and practical way to go. However, for some builders, the experimenter's journey is what counts. That is, the fun and learning occurs during designing/building/testing and not just operating home built gear.

When the output is taken from between the shunt capacitor and the crystal per Figure 1A, a much cleaner sine wave is available. The photo above right tells this story. The focus of all of the experiments on this web page is boosting this lower distortion, lower phase noise signal into something useful. To increase the base oscillator output voltage, the VCC was raised to 12 VDC and the BJT emitter circuit was tuned per Figure 1C. Resident on my work bench are a few potentiometers and a 10-254 pF air variable capacitor with short attached leads. These parts are inserted into test circuits to allow tweaking of R or C as desired. Once the desired tweaking is performed, the potentiometer or variable capacitor is removed and measured. The closest fixed value R or C is then substituted as appropriate. In this experiment, the highest output voltage occurred when the variable capacitor measured 181.7 pF. Thus in my version, 33 pF and 150 pF capacitors were placed in parallel and are shown later in Figure 3.

The Figure 1A output transformer circuit was again used and the power output was 6.8 dBm. Being tuned to 1 crystal frequency is the biggest drawback of the 1C circuit. Tuning a crystal oscillator as described earlier is easy to do however.

The next task was to design and build a buffer/amplifier. To match the low impedance of the Figure 1C crystal oscillator, a common base amp was built. The circuit was morphed over time, however, the initial design is shown as Figure 1B. It was interesting to note that a series resistor (RX) is required to keep the waveform pure. Any RX value less than 470 ohms compromised the sine wave purity. The 560 ohm R shown was perfect, however, as expected, attenuated the oscillator signal. In order to get a decent output voltage, the common base amp had to be run at 5 mA or greater current and ultimately collector tuning was added to try and realize an output voltage greater than 4 dBm.

Through experimentation I learned that adjusting Fig 1B's tuning, emitter current and RX value all could distort the output of the main oscillator at certain values or settings. Running high current also invites parasitic oscillations and soon it was realized that common base was perhaps not the best choice (at least for me) as another separate amplifier stage would be required to get a decent output voltage with or without an attenuator pad.. After trying a number of different buffer-amplifiers including a 50 ohm feedback amp, I chose a favorite circuit which I know has excellent gain plus back to front isolation and would not distort the oscillator waveform; a lightly coupled JFET amp.

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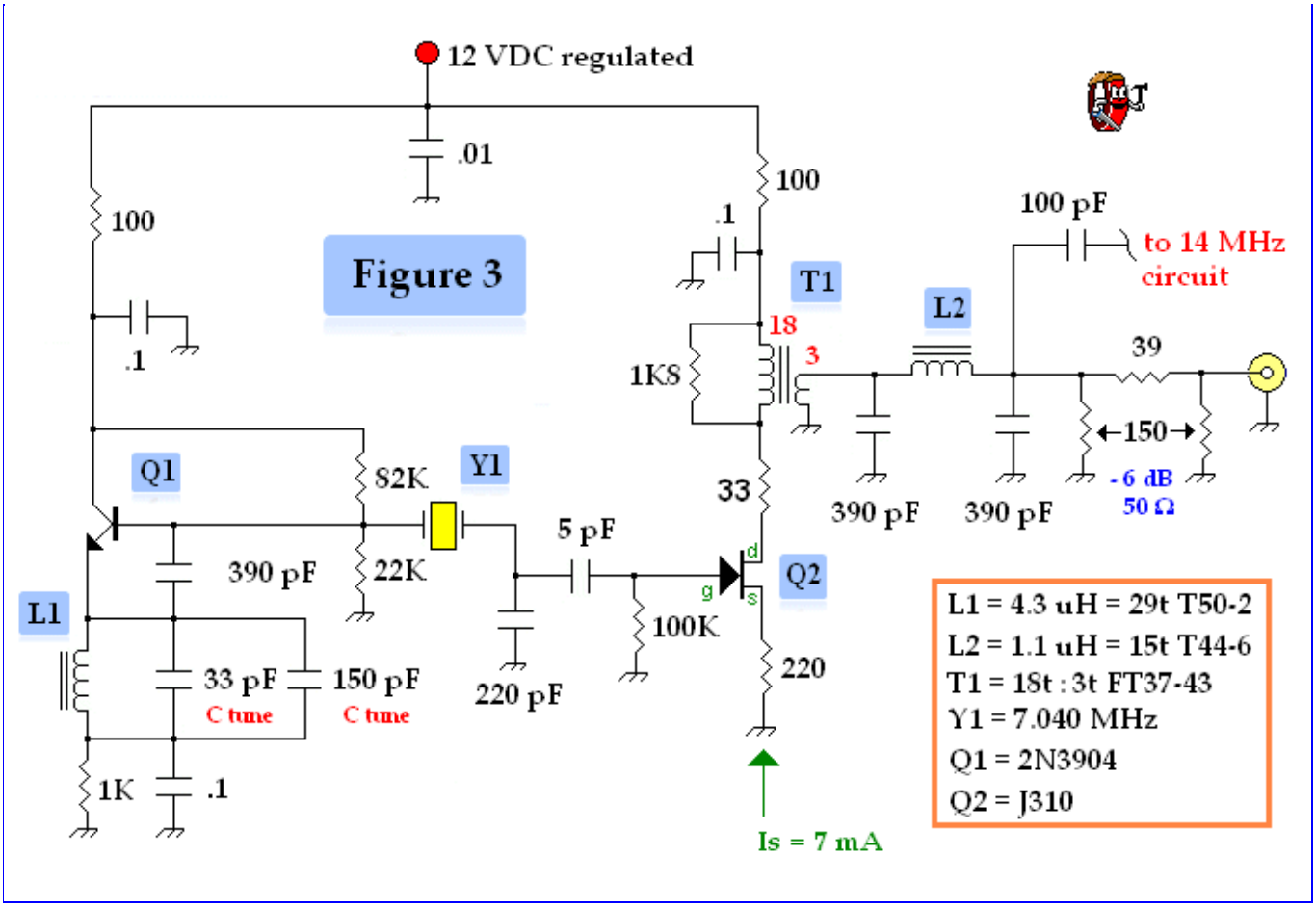
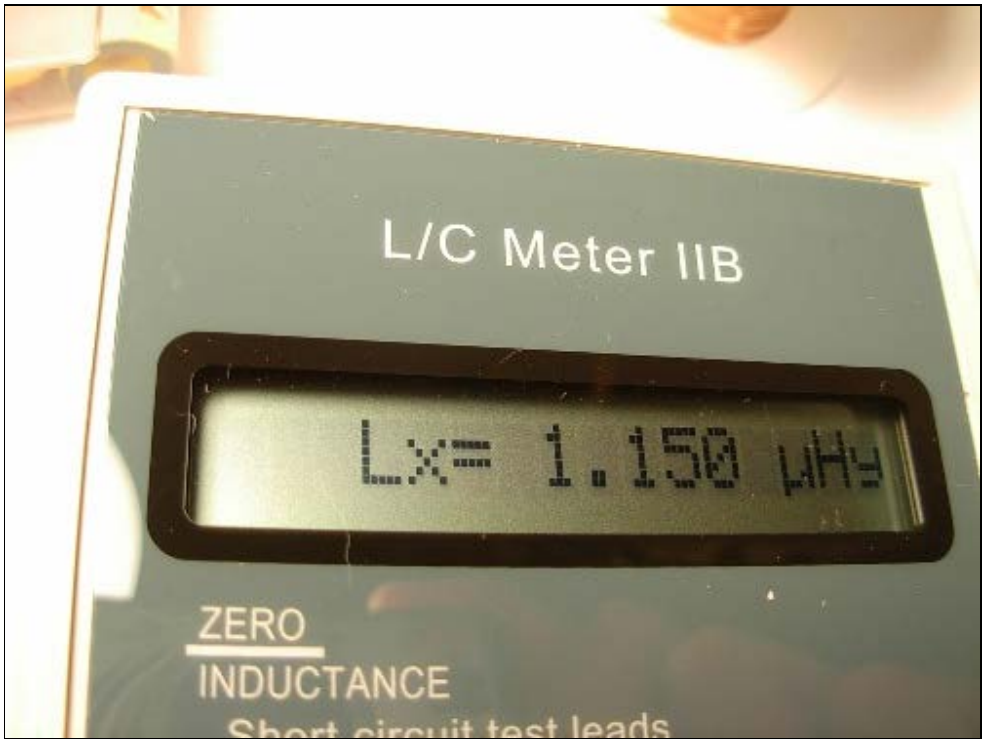
7 MHz Circuit

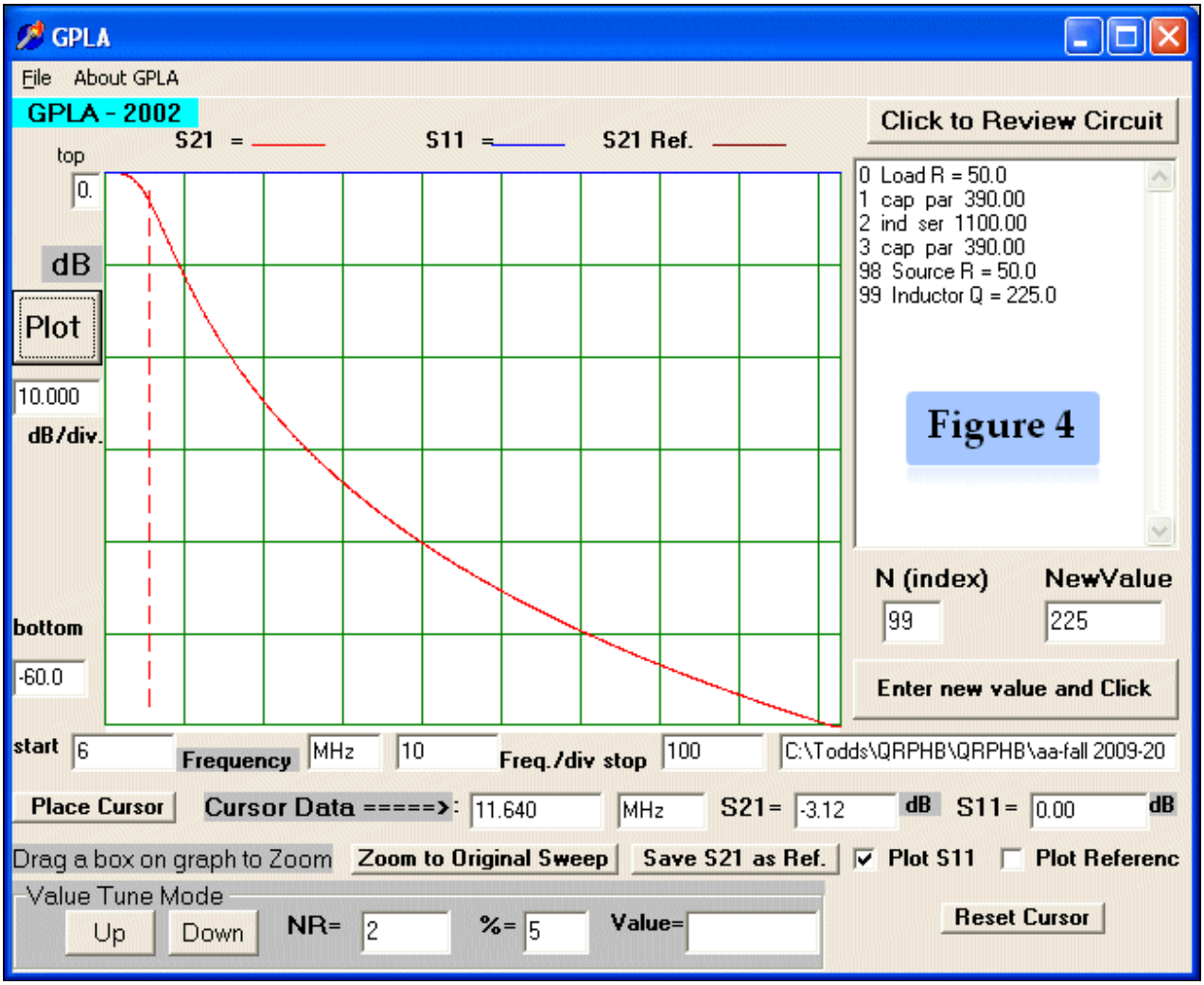


In Figure 3 is the complete schematic of the 7 MHz portion of the low distortion crystal oscillator. The output was 0.52 dBm. A 6 dB 50 ohm pad ensures a well-buffered 50 ohm termination. This aids in calculating gain or loss in circuits it drives. You could easily decrease this to a -3 or -4 dB pad. This reflects the wisdom handed down by our mentors who encourage building RF stages in 50 ohm impedance blocks. A Q2 source bypass capacitor was not placed as it increased harmonic distortion in the output signal. The output is filtered with a simple pi filter. The 100 pF coupling capacitor connect this circuit to Figure 6; the 14 MHz circuit.

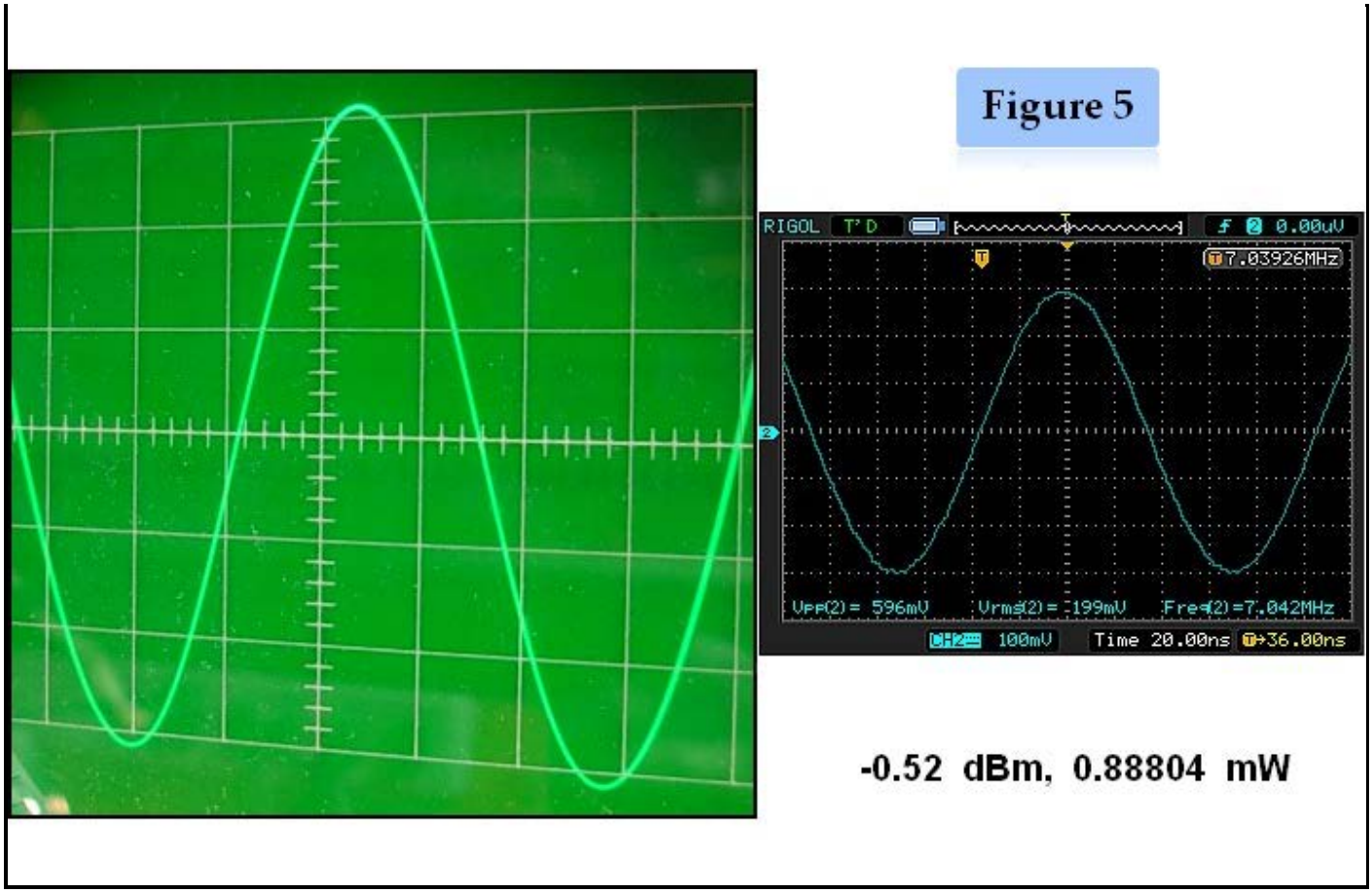


L2 was wound and measured. It is desirable (but not absolutely necessary) to perform measurement on powdered iron toroids to compensate for

variations in wire spacing and toroid permeability. I used a T44-6 core; use whatever appropriate powdered iron toroid you want.



A GPLA analysis of the Pi low-pass filter is shown in Figure 4. The basic circuit was designed with PI Filter Designer on this page and tweaked in GPLA. You may wish to omit this filter or perhaps, design a better one yourself.



The 7 MHz output waveform on the Tektronix (left) and the Rigol oscilloscopes. On the Tek scope, the output power was 0.50 dBm and on the Rigol scope it was 0.52 dBm. The Rigol is an amazing oscilloscope, but only has 256 horizontal lines of resolution, therefore cannot replicate the stellar and beautiful waveform tracings of the ancient Tek scope. In reality, probably, no other modern scope can. I have received many positive comments about the old Tektronix oscilloscope waveforms. It is important to mention, that Rigol waveform viewing is not bad, just very different. The visual display is incredibly accurate and its triggering options, bandwidth, sampling rate and waveform display tools are fantastic. фантастический !

During these experiments, the 40 year old (plus) Tektronix scope was distorting frequencies greater than 10 MHz and breaking into oscillations. After 3 major repairs in 2009, the scope replaced with a Rigol DS1052E. Signal viewing will certainly different; that is for sure. The decision to move from a cathode ray oscilloscope (CRO) to a digital storage scope (DSO) was not taken lightly.

DSO versus CRO - some comments from the workbench

Choosing a CRO versus a DSO is an individualized process. It is your decision alone. Questions to ask yourself may include: What are my needs? What is my budget? Do I have weight and/or space constraints? Carefully weigh the advantages and disadvantages of each.

Proponents of CROs state that these scopes cannot generate artifacts, nor distort the signal. This of course is true as long as the scope bandwidth is adequate. Further, some people feel that aliasing or artifact generated in DSOs due to undersampling (taking too few samples of a waveform) is unacceptable. They may even feel that DSOs are not precision measurement instruments as a result. Limited horizontal screen resolution in DSOs is also a bugaboo for some experimenters and provides further evidence of DSO inferiority in the minds of these folks. These concerns are indeed valid; however, black and white thinking is a little out of fashion in a world more containing shades-of-gray.

The DSO takes a series of samples and stores them in memory. When sufficient samples are present, they are assembled and displayed. The sampling rate of a DSO is variable and depends on the time base setting used. Modern DSOs, like the Rigol, Tektronix Oscilloscopes and Agilent Oscilloscopes have better sampling rates and larger memories than their predecessors and hence aliasing is less of a problem than before; although in some measurement situations, undersampling can occur. One must always well consider and interpret whatever you are measuring and if you use a DSO, always keep undersampling in mind. When first using a DSO, you are on the bottom of a learning curve, however, with attentiveness and practice, one can learn to look for and possibly mitigate undersampling should it occur. In certain cases, a CRO will be superior to a DSO. In my discussions with others about Rigol signal viewing , only 1 significant "aliasing" problem has been noted by a builder when he tested a balanced modulator. The display did not give the expected result (was not filled in as expected) and a CRO was pulled out and the problem was verified. The builder knew there was a problem and could understand why it occurred. This builder also wrote that this was not so much a problem, as a reality of using a DSO.

Some techniques I have gleaned from the Internet about detecting aliasing may include the following:

Vary the time base over several ranges. Events occurring near the time base should be reproducible and if they are not, undersampling might be occurring.

As possible, use a single sweep and dot display. The 'dots' will indicate just where the scope took each sample. If the dots are far apart relative to the waveform timing, aliasing is a possibility.

Some techniques to minimize aliasing:

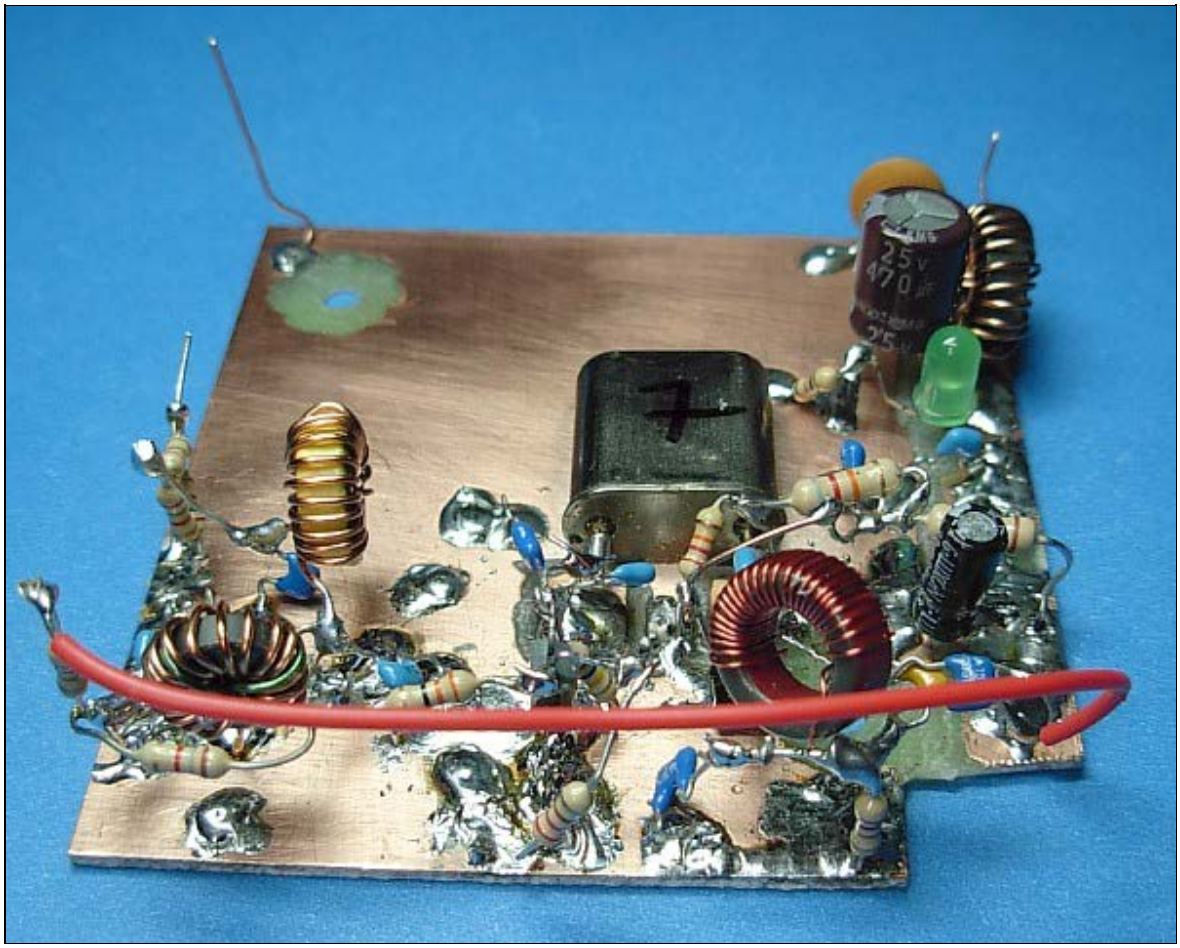
Choose linear interpolation when using math functions.

Use bandwidth limiting in low level measurements (The Rigol seems to automatically use B/W limiting in these situations).

Use trace averaging for low level measurements as possible.

The Rigol weighs 50 times less than my old scope and fits on a small shelf in my small workspace. As a hobbyist, it meets my needs and budget plus has some very cool features. Undersampling is considered and in some cases, such as low level measurement, an analog scope might be a better choice. Happily, a CRO is available to me if I really stress out over it. Signal viewing was taken for granted with my old scope. In some ways, this DSO has prompted me to dig deeper; to become more vigilant and thoughtful about signal measurement and display. If you have a spectrum analyzer, and use a DSO for signal viewing, the ability to perform slow sweeps while maintaining a perfect display is quite enjoyable. Again, please decide the CRO versus DSO issue for yourself (or maybe get a hybrid). The DSO is not a perfect solution to every signal viewing situation, but their constraints are quite livable considering their numerous modern features.

DSO's: “They are not your father’s oscilloscope”; that is certain!



The Figure 3 breadboard. AWG 24 to 26 gauge wire was used in the various inductors and transformers to better secure or anchor these parts.

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14 MHz Circuit

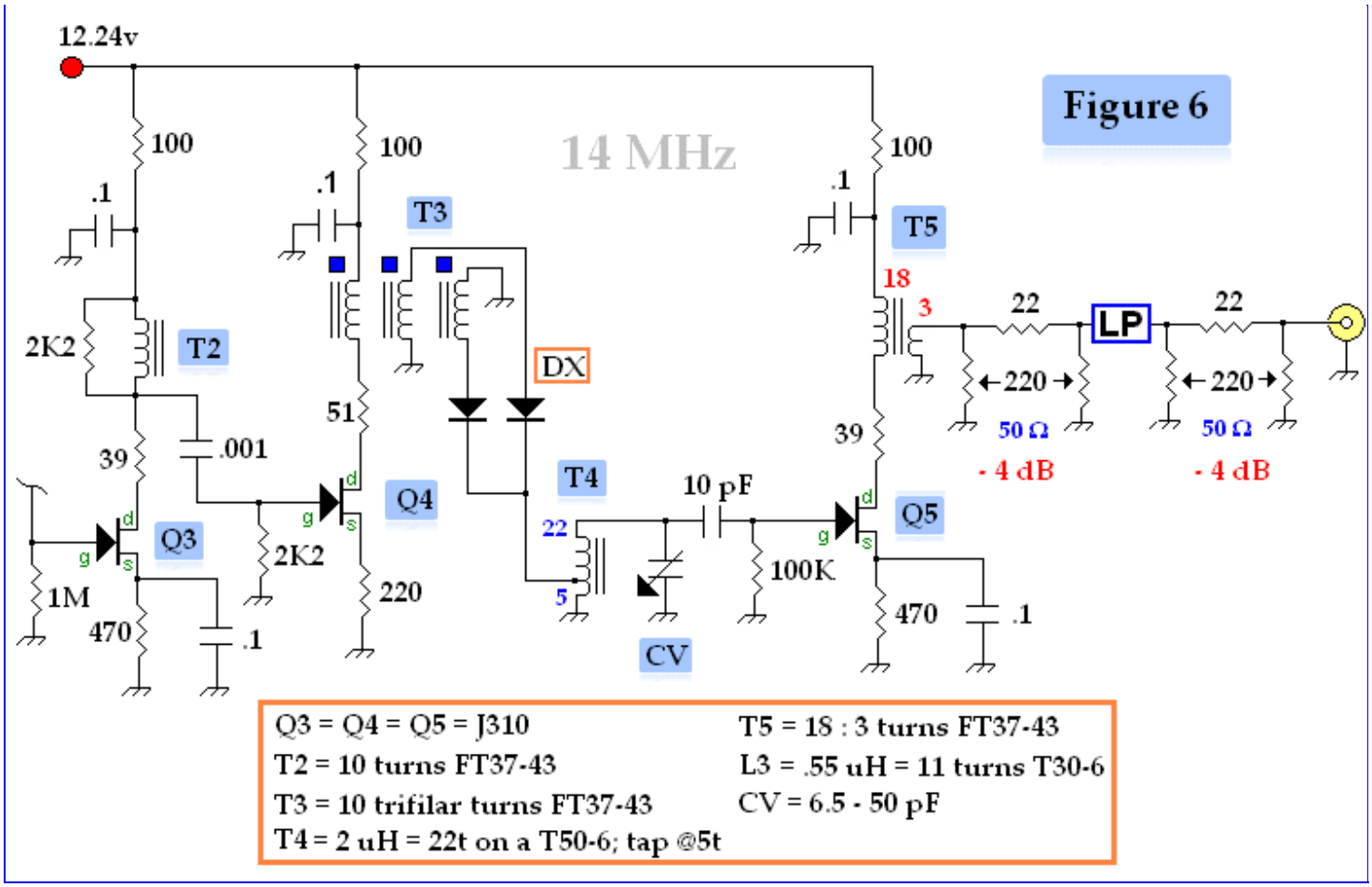


Figure 6 is the final schematic for the 14 MHz circuit as developed on the workbench. Your design might look very different than mine. In the first version of this circuit, there were only 2 JFET amplifiers and the resultant output voltage was too low (even without a 50 ohm attenuation pad). To compensate, I ran the source current of the JFETs above 15 mA, placed a source bypass capacitor on Q4 and also used a 1000 pF capacitor to couple the input to Figure 2. Some fairly bad harmonic distortion was measured at the output and it seemed crazy to run so much current. Therefore it was decided to run a third JFET amplifier and use only modest current in the trio of JFET amplifiers.

CV tunes very sharply and required some care when peaking the output voltage.

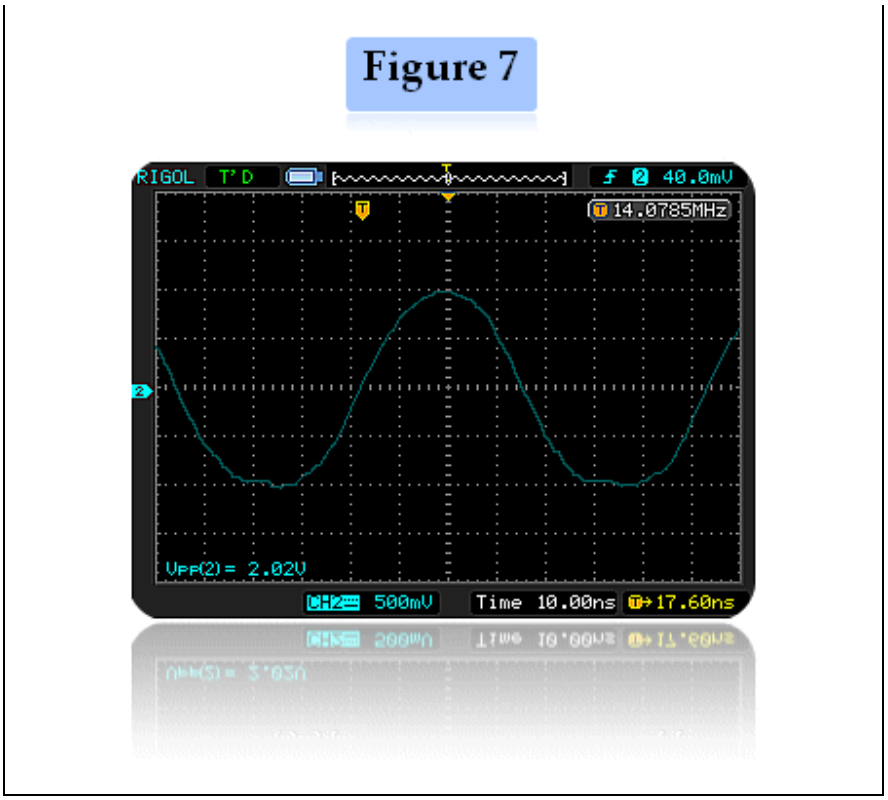
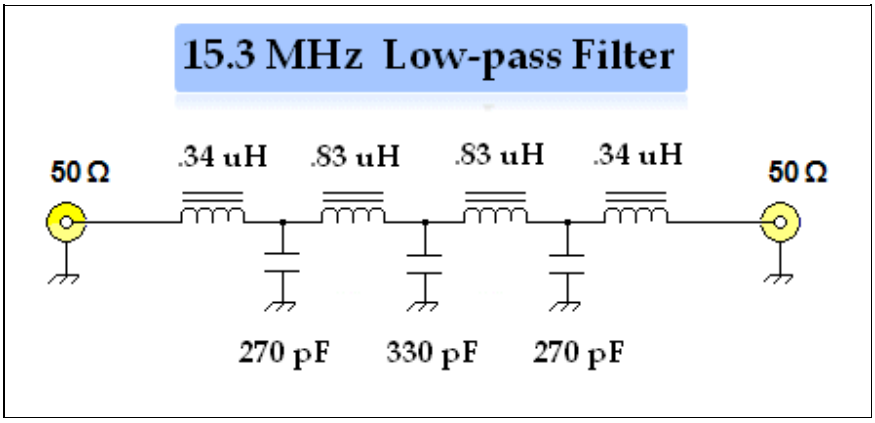


Figure 7 is a screen capture of the 14 MHz circuit output measured using a sensitive 50 ohm terminated oscilloscope. This was with no low-pass filtering. The output is distorted. Presumably this happened in the diode frequency doubler. This is not a low-distortion oscillator.



A N = 7 Chebyshev low-pass filter was inserted in Figure 6 at point LP. I checked with a spectrum analyzer and the 2nd harmonic was down 38 dB. There were no other measurable harmonics after that.

Figure 7b

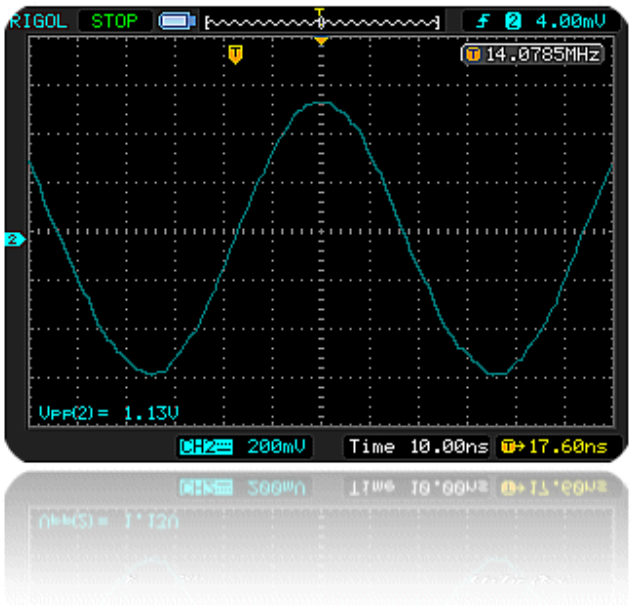


Figure 7b is a screen capture of the 14 MHz circuit output after the low-pass filter and final attenuator pad. Vpp on this graphic = peak to peak voltage = 1.13 volts. The output power is 5.04 dBm, or 3.19 mW . The Q4 and Q5 source resistors and the output attenuator pads are 2 areas of the circuit where you might easily change the output power. In the end, the circuit labeled Figure 6 was chosen. Your output voltage will probably vary, but can be easily adjusted as described.

A

Figure 8

B

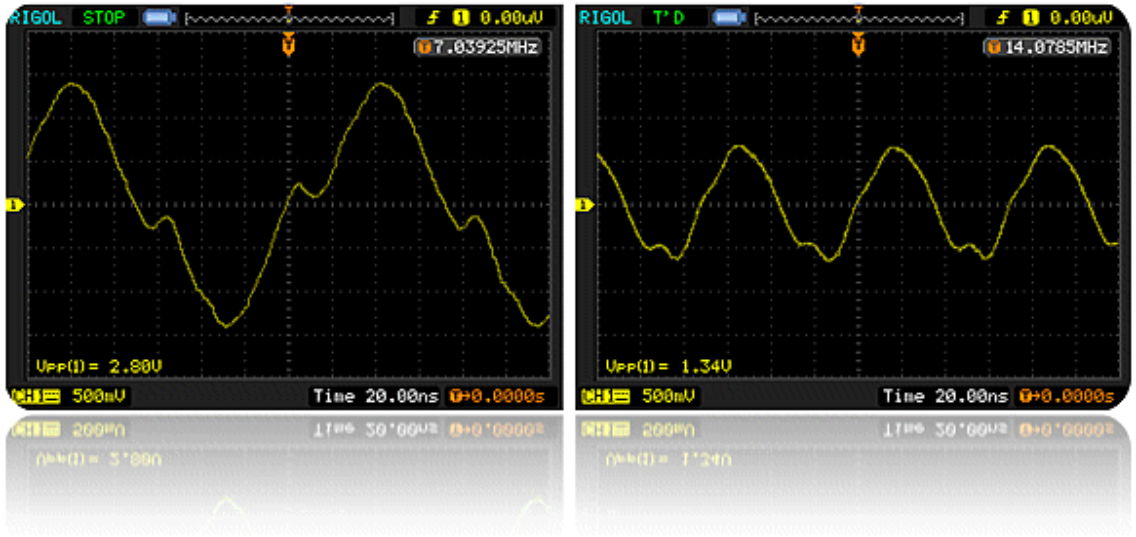
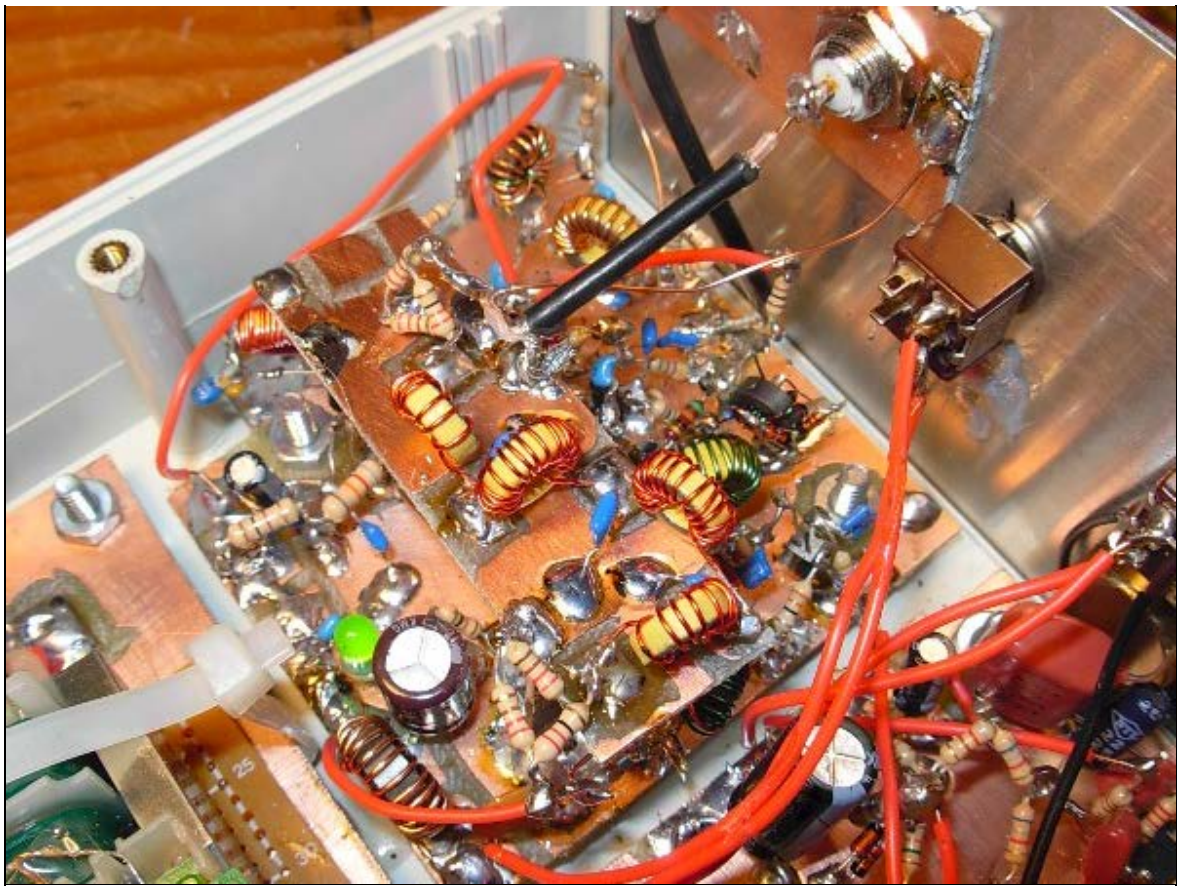
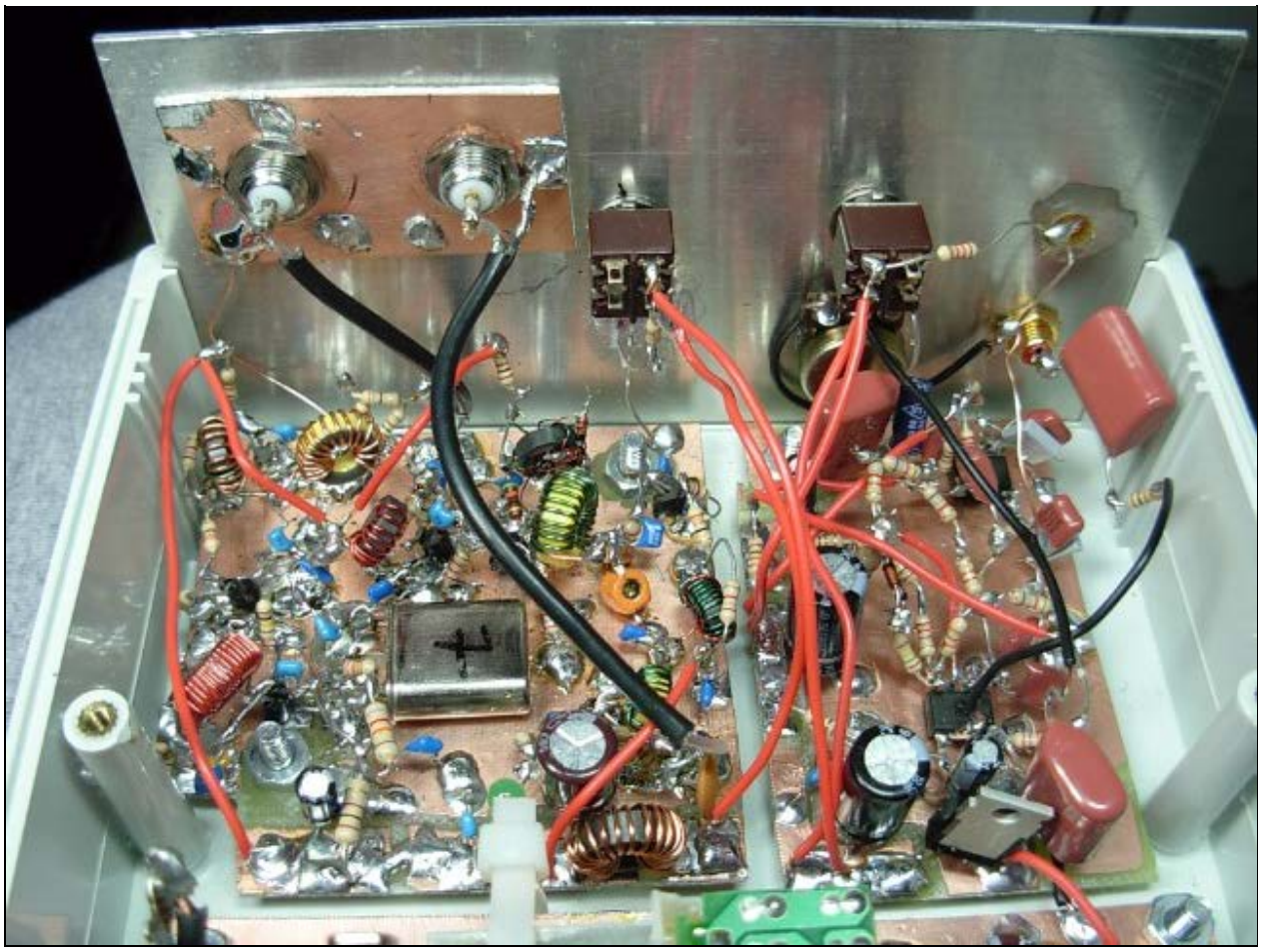


Figure 8 are Rigol digital screen captures. Figure 8A is a measurement taken from the anode at point DX from Figure 6. Figure 8B is a measurement taken at the cathode of DX and shows distortion caused by the diodes, reduced AC voltage and of course, frequency doubling.



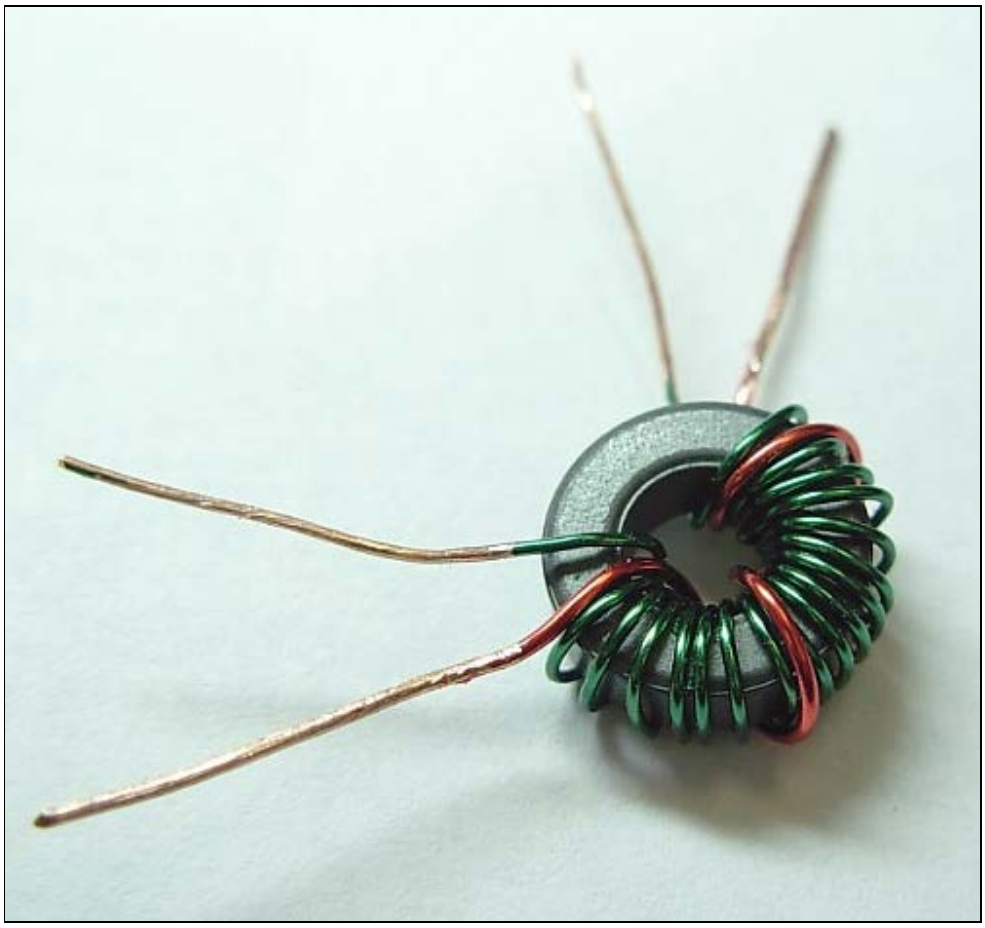
The 7 and 14 MHz circuits bread boarded and mounted in a chassis. The 7.039 MHz and 14.079 MHz outputs are connected to BNC jacks via RG-174 cable. The 15.3 MHz Chebyshev low-pass filter and -4 dB attenuator pads are on a raised Ugly Construction board. Some VCC decoupling parts are also on the bottom RF board.



The 7 and14 MHz crystal oscillator board mounted in a project chassis along with a 1 KHz low noise oscillator This photo was a prototype version that did not have a Chebyshev low-pass filter after the 14 MHz stage.

Front view of the .001, 7 & 14 MHz oscillator. It is really fun to build your own test equipment.

Transformer Notes and Conclusion



A photo of T5 from Figure 6. Some builders have emailed and stated they do not like to wind inductors/ transformers. I always ask them why? Often these builders were concerned with little details such as wire gauge and spacing, choosing the core size and which magnetic material to use. The Radio Amateur literature is replete with great tutorials on winding coils using toroids. Truly; the more toroids you wind, the easier it gets. Here are some simple points for beginners:

Powdered iron toroids are generally for tuned circuits. I.e. A capacitor and the inductor are tuned to a center frequency. Powdered irons containing the #2 and #6 material tend to tune sharply and have fairly high Q

Ferrites toroids are generally for use in broadband or wideband (untuned) applications. #43 material is relatively low Q and lossy as compared to the number #2 and #6 powdered iron toroids

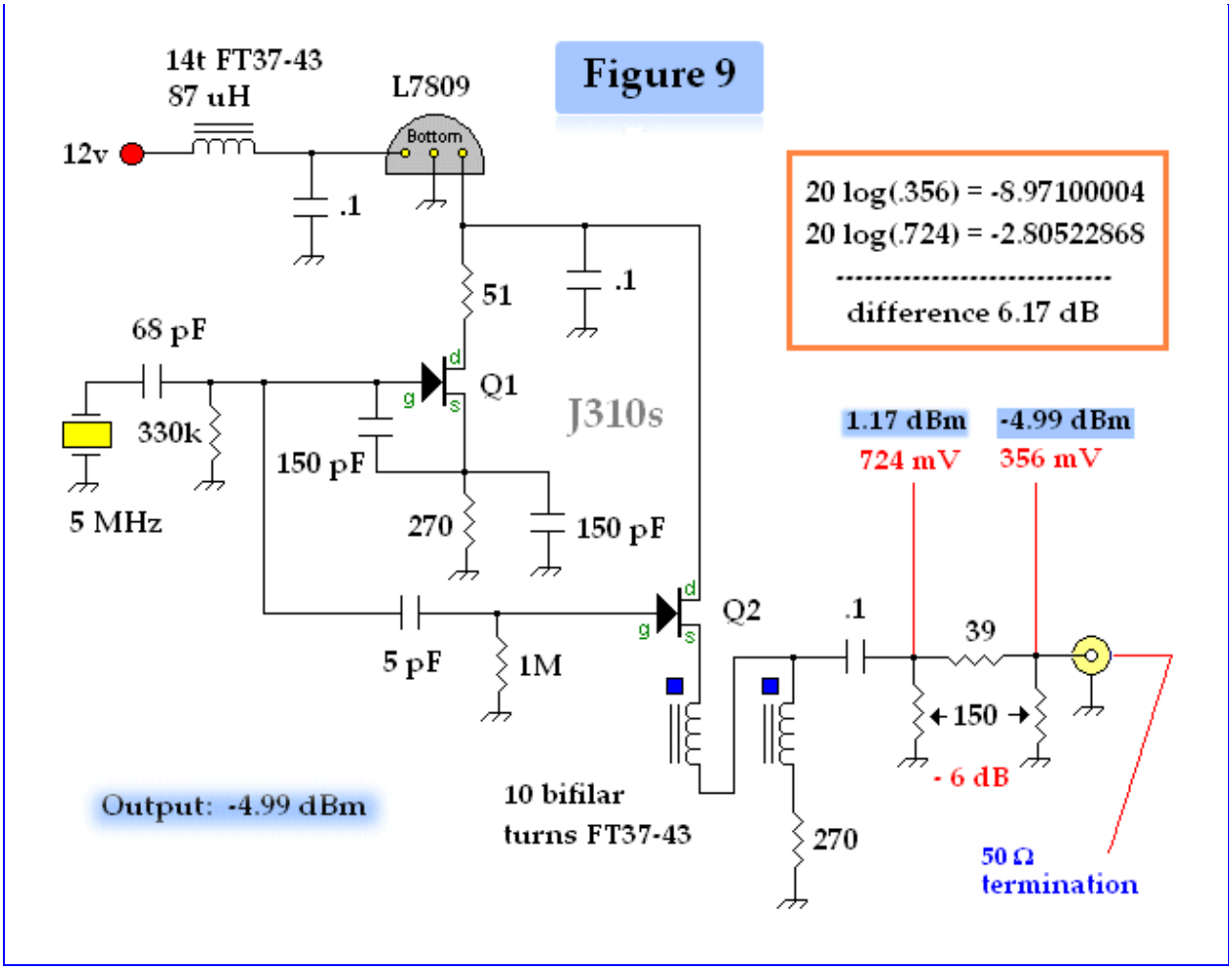
Wind your inductors with enamel coated magnet wire. Popular gauges include 28, 26 and 24, but this is quite variable.

Minimally, you could get by with just #43 ferrite and #6 powdered iron toroids. For example, FT37-43 ferrites and T50-6 plus T68-6 powdered iron toroids could build a lot of inductors/transformers. In the photo above, I used #24 AWG wire for the 18 turns and #22 AWG wire for the 3-turn link. The 3 turn link is grounded on one end and well anchors the transformer. Thicker wire was chosen because Ugly Construction was used and the part is really anchored with the #22 AWG wire.

Conclusion

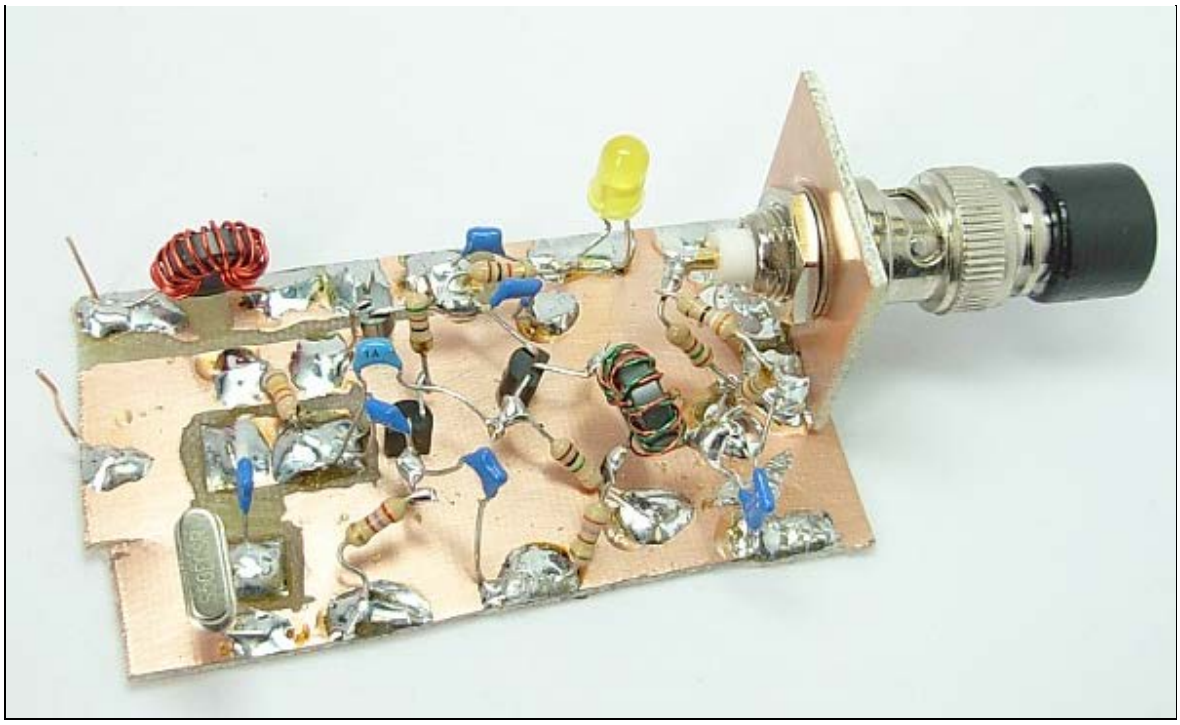
It took quite a lot of experimentation to reach the project goal. The experience was pleasant and comparing the old and new oscilloscopes was an added bonus. Perhaps, the circuits are overly complicated, however, they are critical signal generators for my workbench. Well designed signal generators have extensive RF-proof shielding to allow serious attenuation. These will have to do.

Addition - Jan 31, 2010 5 MHz JFET Low Noise Oscillator



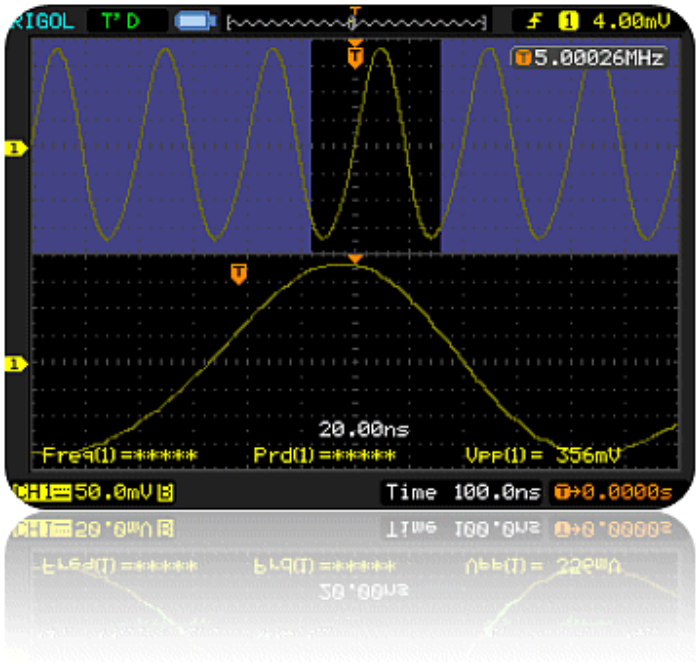
Another experimental low noise oscillator was built for 5 MHz. The desired output power was -5 dBm. The schematic and peak-peak output voltages and powers are shown in Figure 9. The 68 pF capacitor in series with the crystal was found experimentally by using a variable capacitor and measuring the output voltage and observing the scope waveform. When these appeared to be optimal, the trimmer capacitor was removed and measured at 67.17 pF. This circuit can easily be adjusted to give higher output power such as 7 dBm. To increase power, you may consider increasing the VCC to 12 volts, add a 0.1 uF source bypass capacitor in parallel with the 270 Ohm resistor of Q2 and/or adjust the pad attenuation. There are other circuit alterations to increase power, but the aforementioned are a good start.

It is important to measure your output voltages with a 50 ohm load connected to the device. Standard value resistors were used which throw off the value somewhat, but the actual attenuation of the pad is very close to the 6 dB attenuation expected. The math can be done with software such as Applets H and I on the QRP tools web page.

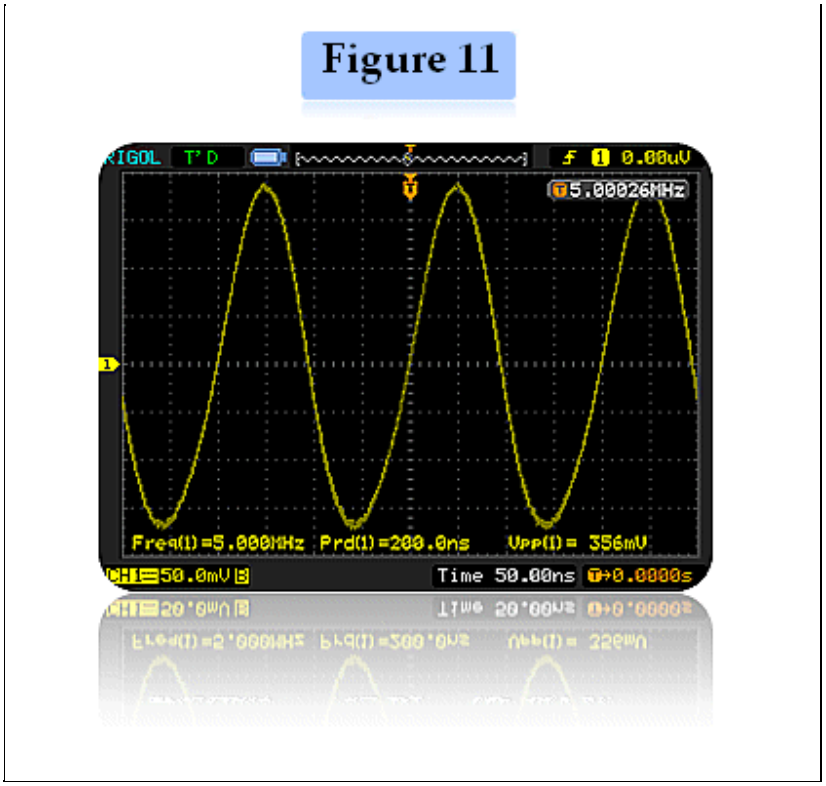


The breadboard of Figure 9 is shown above. The - 5 dBm output was required for some upcoming experiments and to study log and power measurements. The black BNC appliance is a 50 ohm load. These are very handy, but not a necessity. In most cases, a simple 51 ohm resistor to ground is the load.

Figure 10

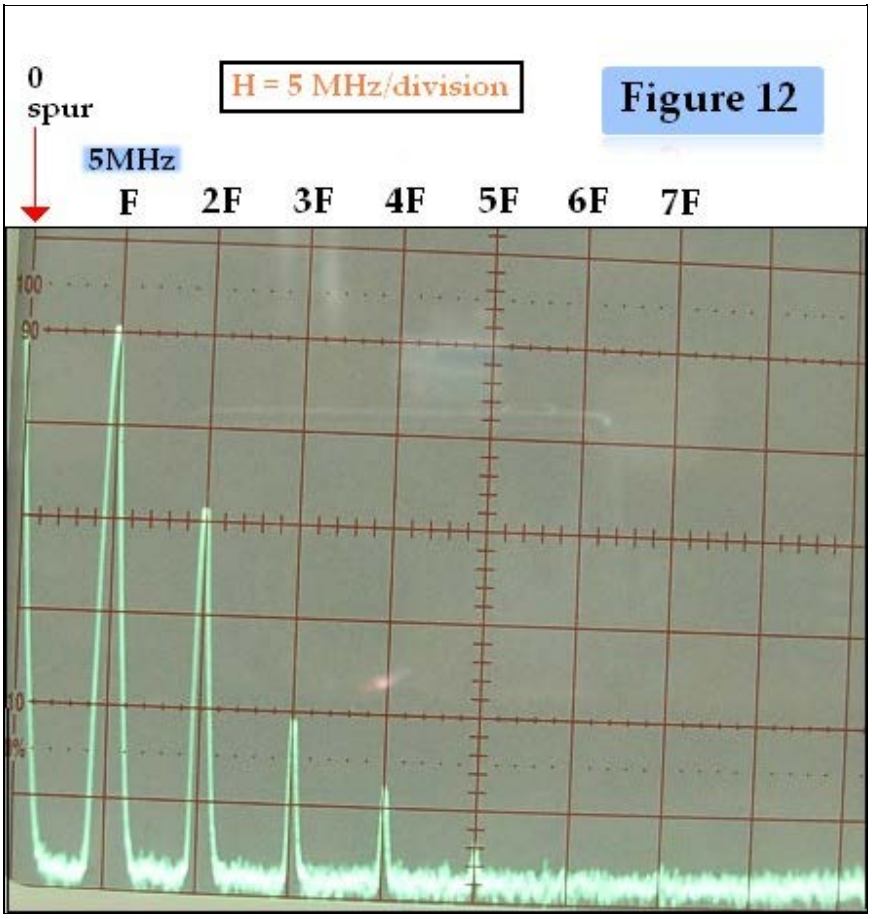


A screen capture from Figure 9 signal viewing maneuvers.



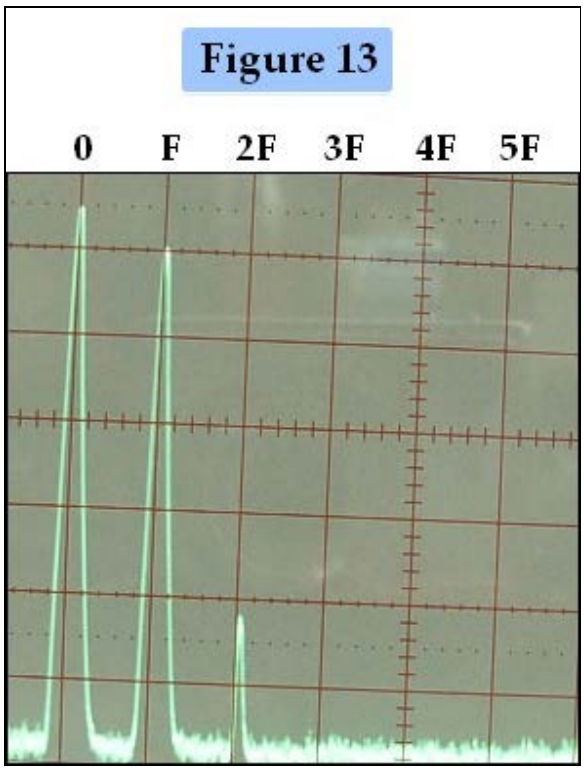
The output waveform looks good - even on a DSO!

Spectrum Analysis

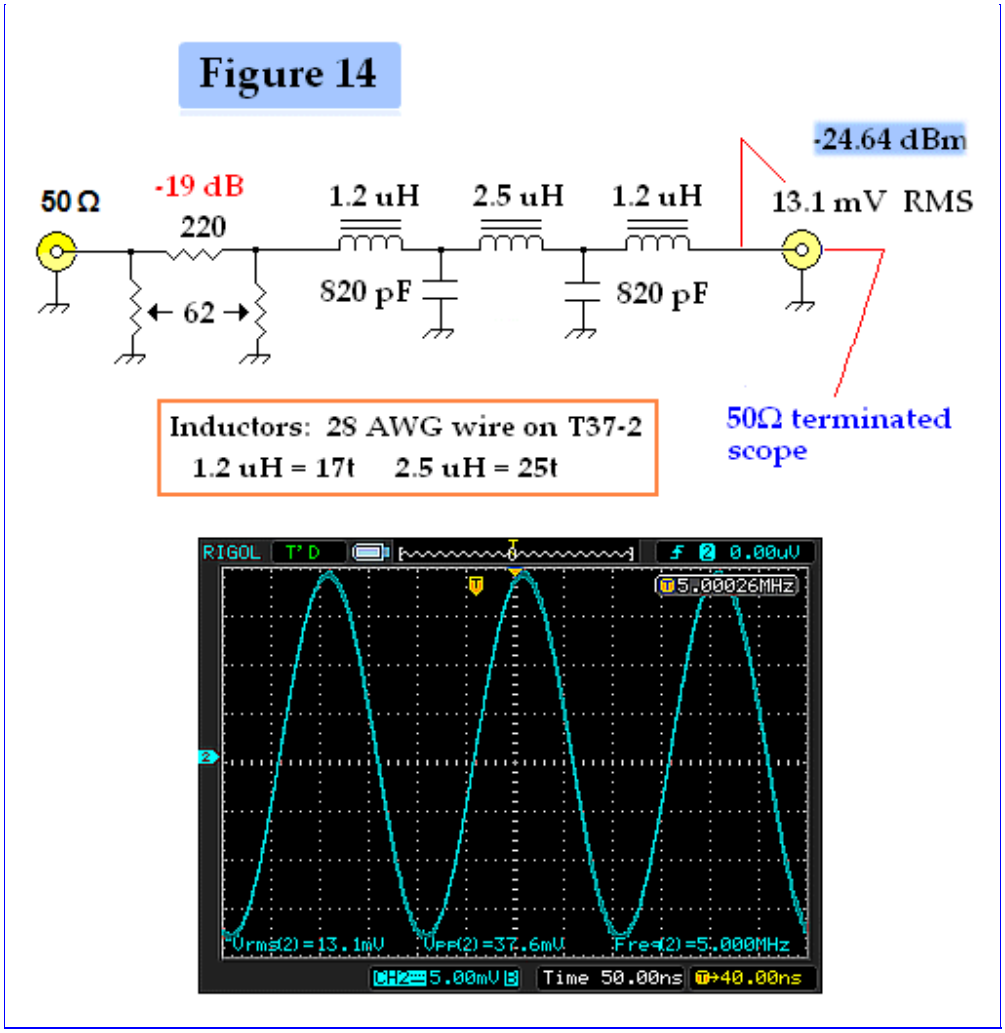


The output of the Figure 9 oscillator was further attenuated -19 dB for analysis in a spectrum analyzer. The -19 dB was chosen as this 50 ohm pad uses near standard resistor values and would well ensure a very safe signal level for the spectrum analyzer. The pad is shown in the Figure 14 schematic. The second harmonic (2F) is down about 20 dB from the fundamental. Each horizontal division is 5 MHz and each vertical division

is 10 dB. Compared to other more conventional oscillators that were checked in the spectrum analyzer, this oscillator has low harmonic content.



A N=5 or 5 element Chebyshev low-pass filter was placed after the Figure 9 oscillator and connected to the spectrum analyzer to see what happens. The 3 dB down frequency of this low-pass filter was 6.53 MHz. The second harmonic is now ~42 dB down. There are no measureable harmonics after 2F. It is really cool to "see" what a filter does to a signal.



The schematic of the -19 dB attenuator pad and the low-pass filter is shown. To measure the output, the 10X probe was disconnected. The circuit was connected to the oscilloscope via 50 ohm coaxial cable and the scope input was terminated with 2 paralleled 100 ohm resistors. The 50 ohm scope termination technique will be discussed in a future web page addition. The RMS voltage values were inputted into Applet J on the QRP Tools page to calculate the output power. The RMS output was 0.131 volts which calculated to an output power of -24.6 dBm.

Crystal Parameter Checker

Introduction

This web page is a supplement to JavaScript Applet **G** on this web page. This software does the math using a simpified version of the method to determine motional inductance and capacitance developed by David, G3UUR. This is a very basic tutorial meant as an introductory guide for novice builders.

Shown in the photo above is 1 of my crystal parameter checkers. The schematic may be found in many places including EMRFD Figure 3.35 (See Errata) and on this pdf by Nick, WA5BDU. A power indicator LED has been added, but the circuit is the standard design. In this breadboard, the crystal being measured is tack soldered in.

Many builders just copy other builder's I.F. filter schematics, however, your crystal filters will perform better if your design is based upon the exact parameters of the crystals you have. For the simple design or optimization of a crystal filter, it is necessary to measure crystal parallel capacitance plus take other measurements to calculate motional inductance and capacitance. Determining your crystal parameters is not difficult if you have a capacitance meter, a frequency counter and some math skills. It is easiest to use a program to crunch the math; hence I wrote a stupid-simple JavaScript applet.

Designing filters is another story; it takes knowledge, practice and good software for this. Filter design theory has been extensively covered by Anatoly Zverev, Wes Hayward and others. The work of Nick, WA5BDU is also greatly appreciated. His presentations and references are excellent for those keen on learning more about filter design.

A four crystal 5.00 MHz SSB I.F. filter was desired. 20 crystals were on hand — they were from the same batch. The crystals were all placed in the above oscillator and their frequency was measured. The 4 crystals closest matching in frequency were set aside. The crystal parameters of these 4 were then determined. Typically these values are averaged and this average is used to design or tweak the filter using software.

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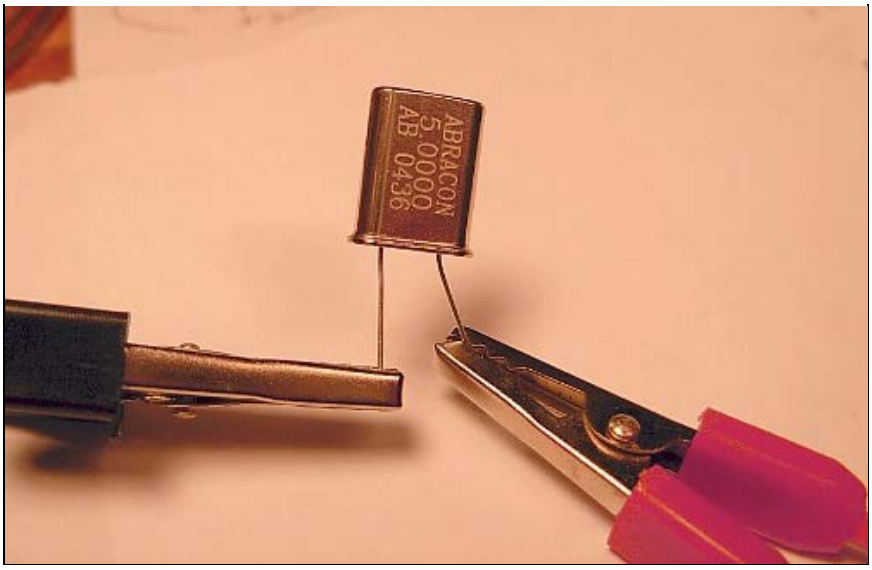
1. Measure Capacitance

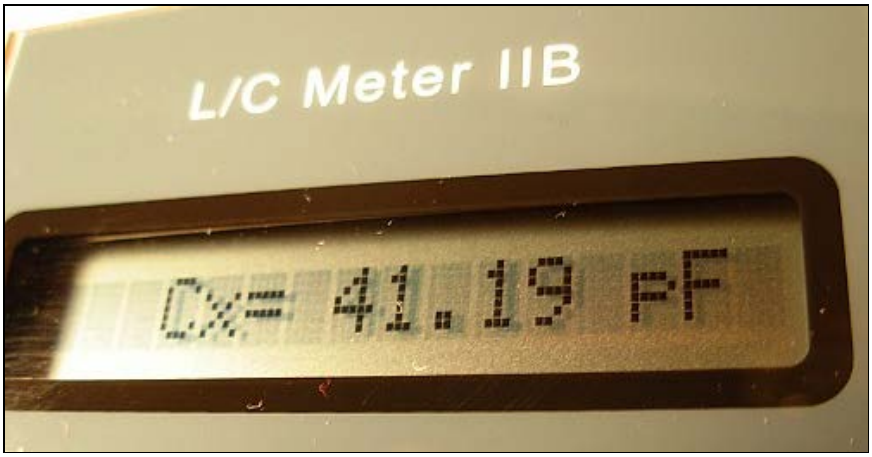
The procedure for determining the parameters of 1 crystal is described.

The first step is to measure the crystal capacitance (called parallel capacitance) using a capacitance meter.



Measuring the crystal parallel capacitance

Parallel capacitance of the above 5.0 MHz crystal in an AADE LC meter



Next, measure the capacitance of the open switch plus the 33 pF fixed value cap wired in-situ. This will give you the total circuit capacitance of the open switch, the 33 pF fixed value capacitor, and any stray capacitance from your crystal holder, wires, etc. The switch itself plus stray wiring will be a few pF so the total should be around 36 to 40 pF or so. In my test oscillator, the result = 41.19 pF as shown. On my other crystal checker with a better switch, it's 36.9 pF.

In the calculation of crystal Lm and Cm, the parallel capacitance and the switch circuit capacitance will be summed.

---

2. Measure Frequencies

A crystal is put in the oscillator with the switch open. Record the frequency. Your counter must have resolution down to 1 Hertz. After recording this value, throw the switch and measure and record this frequency. You now have all the measured values required to calculate motional parameters and adjust or design a filter. Motional parameters are calculated in Applet G.

Frequency measured with the switch open = 4.999274 MHz

Frequency measured with the switch thrown = 4.998317 MHz

---

3. Do Math by Hand or with Software

G. Calculate Lm and Cm For a Crystal using the G3UUR Method

image

Enter frequency in MHz written on crystal (nominal frequency): 5.0

Enter measured frequency in MHz with switch open: 4.999274 Enter measured frequency in MHz with switch thrown: 4.998317

Enter crystal capacitance in pF: 4.65 Enter open switch circuit capacitance in pF: 41.19

Calculate Xtal Parameters Cm = 17.55 femto Farads , Lm = 0.0578 Henries

The applet **G** calculation of the crystal parameters using the above measured values

---

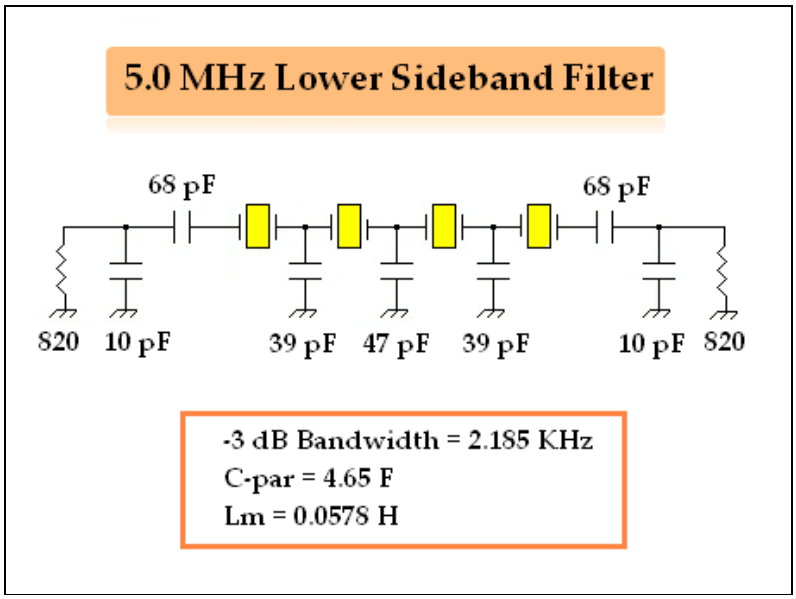
4. Example Filter Adjustment

It is assumed that most builders will use software to design or tweak their crystal filters. The only 2 programs tried *(to date)* include AADE Filter Design and the Ladpac software collection that supplements EMRFD. I am more familiar with the Ladpac programs written by Wes, W7ZOI. Only

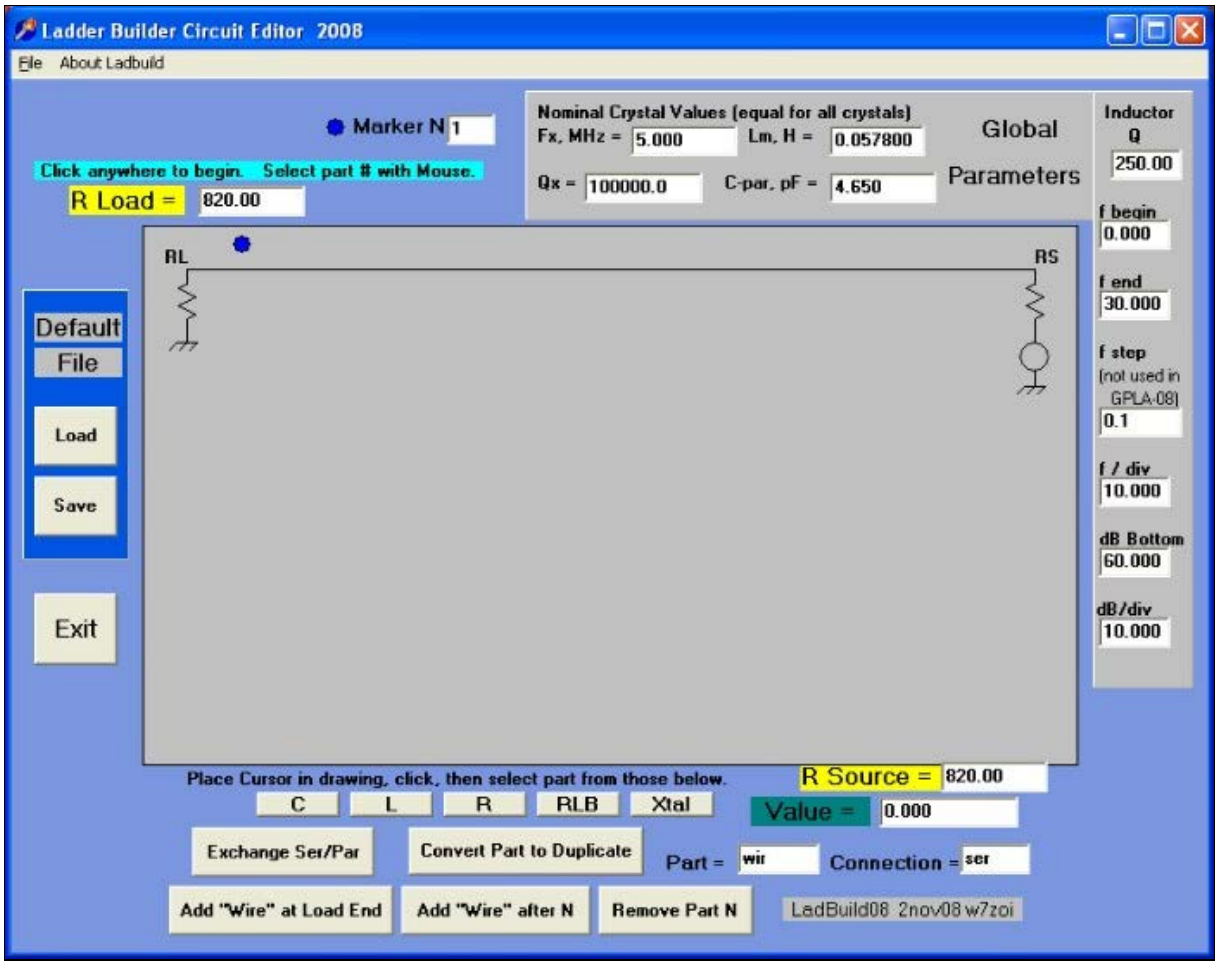
these programs are demonstrated. Please read the instructional file Ladpac2008 Manual.pdf to understand these programs. *The Ladpac software bundle includes GPLA.*

The purpose of this tutorial is not to teach crystal filter design, but to describe a relatively simple method to tweak an existing design using your measured crystal parallel capacitance and its calculated motional inductance.

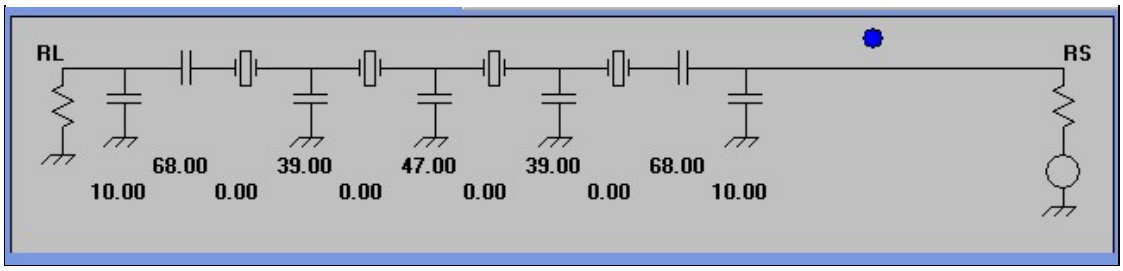
The first step is to digitally format your filter into a file that can be analyzed in GPLA. In my opinion, the easiest way to do this is to use the ladder circuit editor ladbuild02.exe or better yet, its update - ladbuild08.exe. The model filter follows:



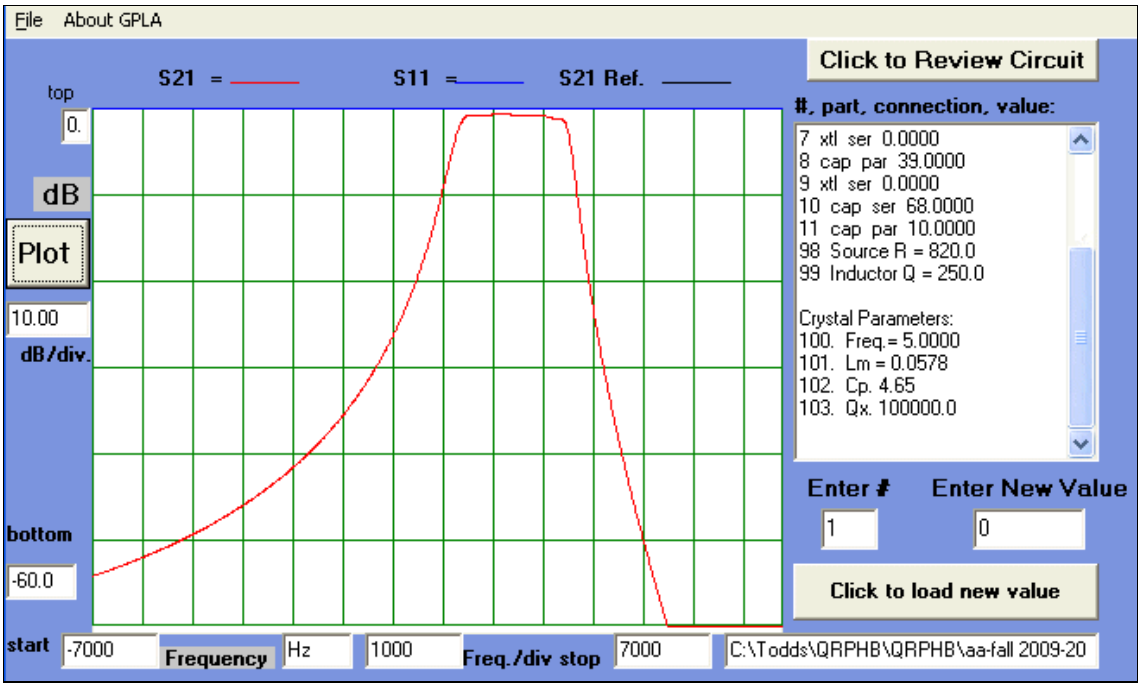
The model 5 MHz SSB filter



Clear any existing components and enter the termination R, C-par and Lm values. Qx is set at 100000

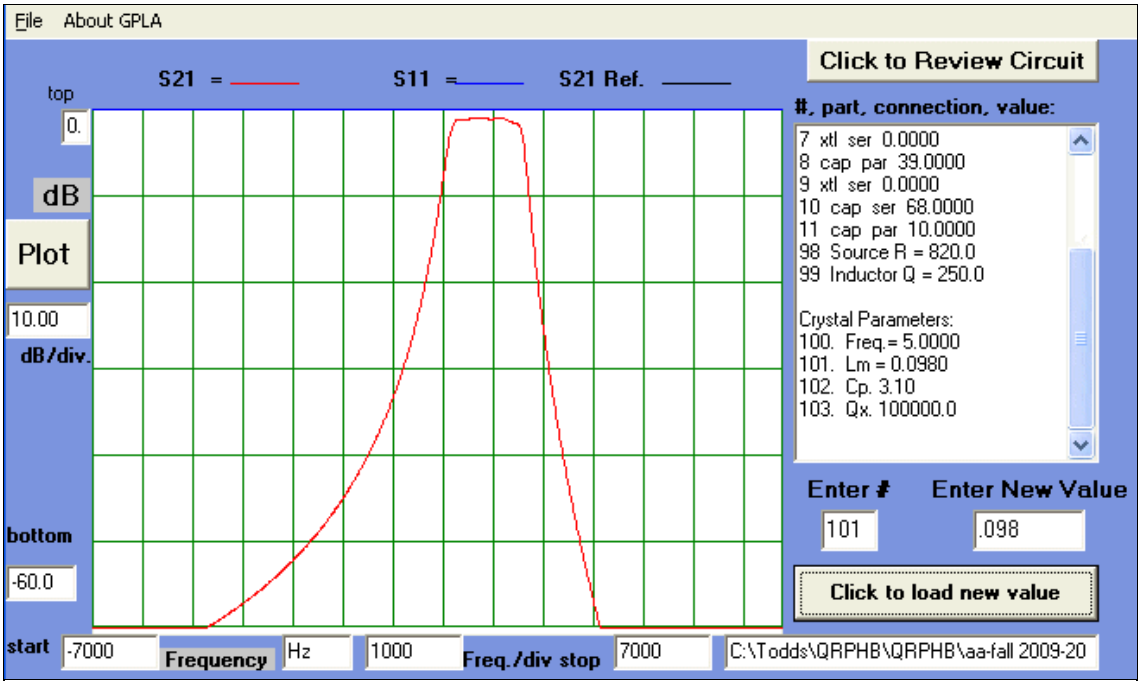


Build your filter within the editor. Save your work. Start up GPLA and load your newly saved filter.



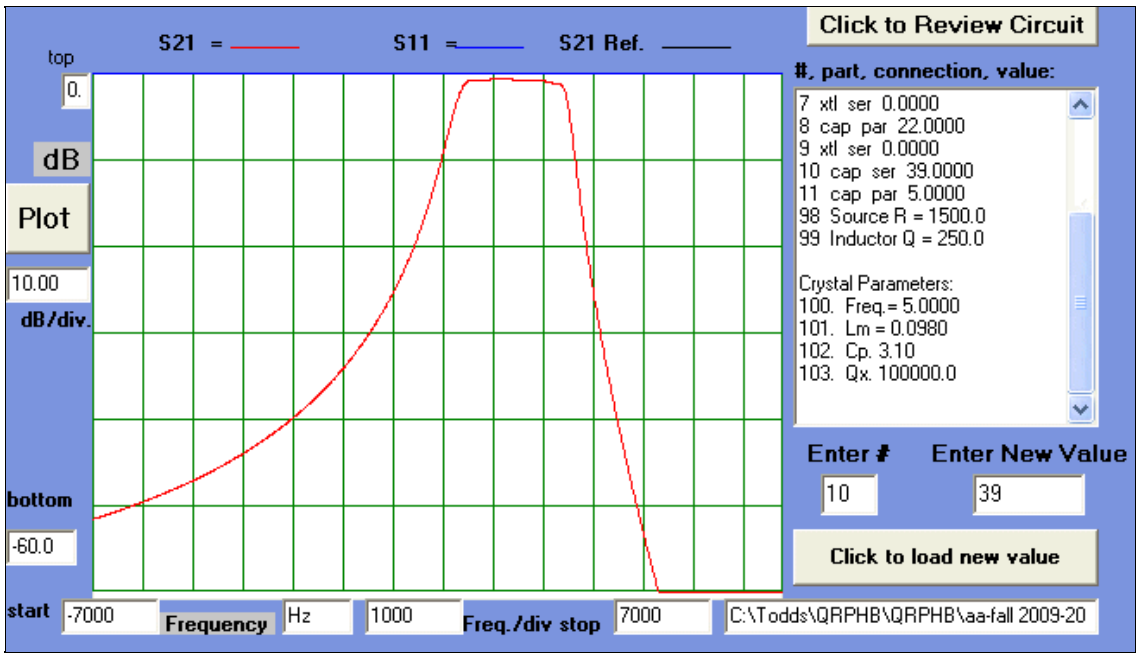
Set a sweep and **x** axis increment (-7000, 1000 and 7000 in this example). Push the Plot button

Let's say you wanted to use this filter design and have determined the average parameters of the 5.00 MHz crystals in your parts collection. Let's assume that for your crystals, C-par = 3.1 pF, and Lm = 0.098H. Input these values in GPLA.



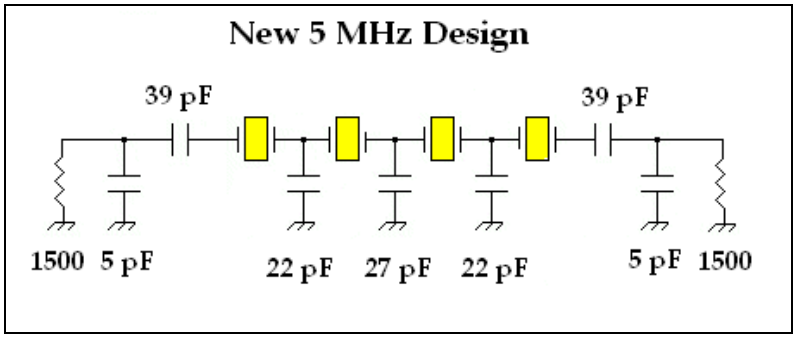
Look what happened to the crystal filter's bandwidth. Our - 3dB bandwidth is now somewhere around 1464 Hertz. This simple experiment illustrates how important it is to use the parameters of your crystals to obtain a desired filter response.

Experiment with the various functions in GPLA to learn how to use it. Set whatever reasonable sweep you want. This program is best learned by using it repeatedly.

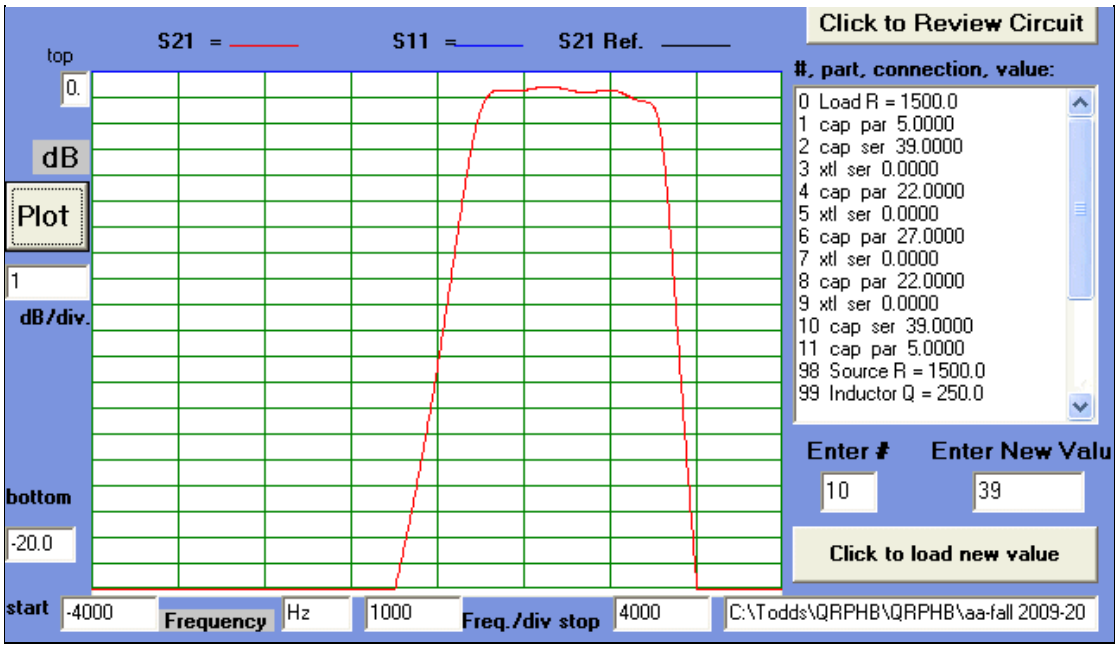


In the above screen capture, the above filter was tweaked to "re-establish" a -3dB bandwidth of ~2.172 KHz. All adjustment was performed entirely in GPLA by swapping capacitor values and observing the resultant waveform. When you get an overall pleasing bandwidth plus shape, but there is too much ripple at the top, generally you must increase the terminating R values. This is the brute-force, manual way to tune filters. For this method, you need not understand terminology such as as series resistance, MESH, K or Q values, Butterworth response, or Chebyshev with 0.1 dB of ripple.

Admittedly, at first, this method can be quite time consuming and tedious, however, with practice, you may be able to tweak a filter in only a few minutes. Clearly, the more you dig into understanding crystal filter design, the better your filters can be, however, getting overly complex can scare off builders who are new to this hobby. Note these filters use standard value capacitors and resistors; perfect for popcorn I.F. filters.



The original 5 MHz Model filter with updated C and R values using Lm = 0.098 and C-par = 3.1



GPLA zoom of the **Y** axis showing the first 20 dB of attenuation.

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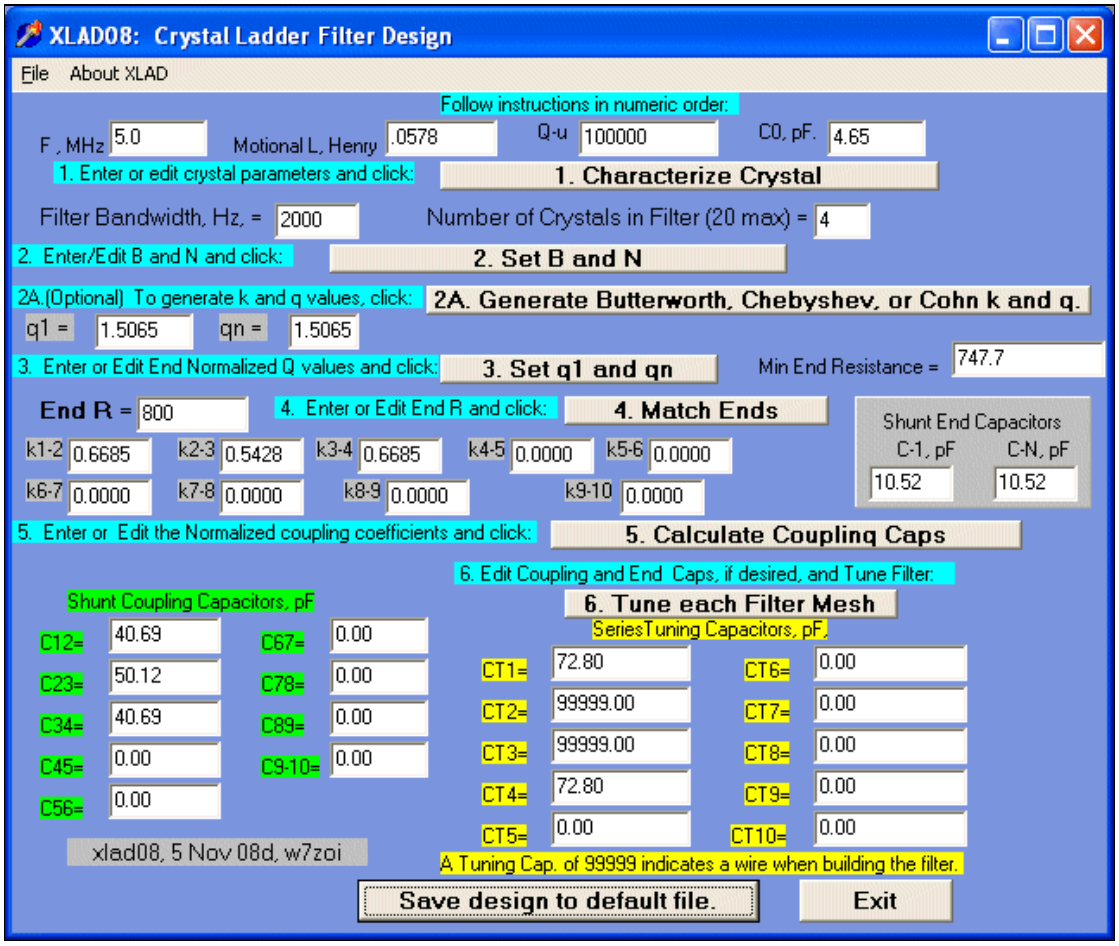
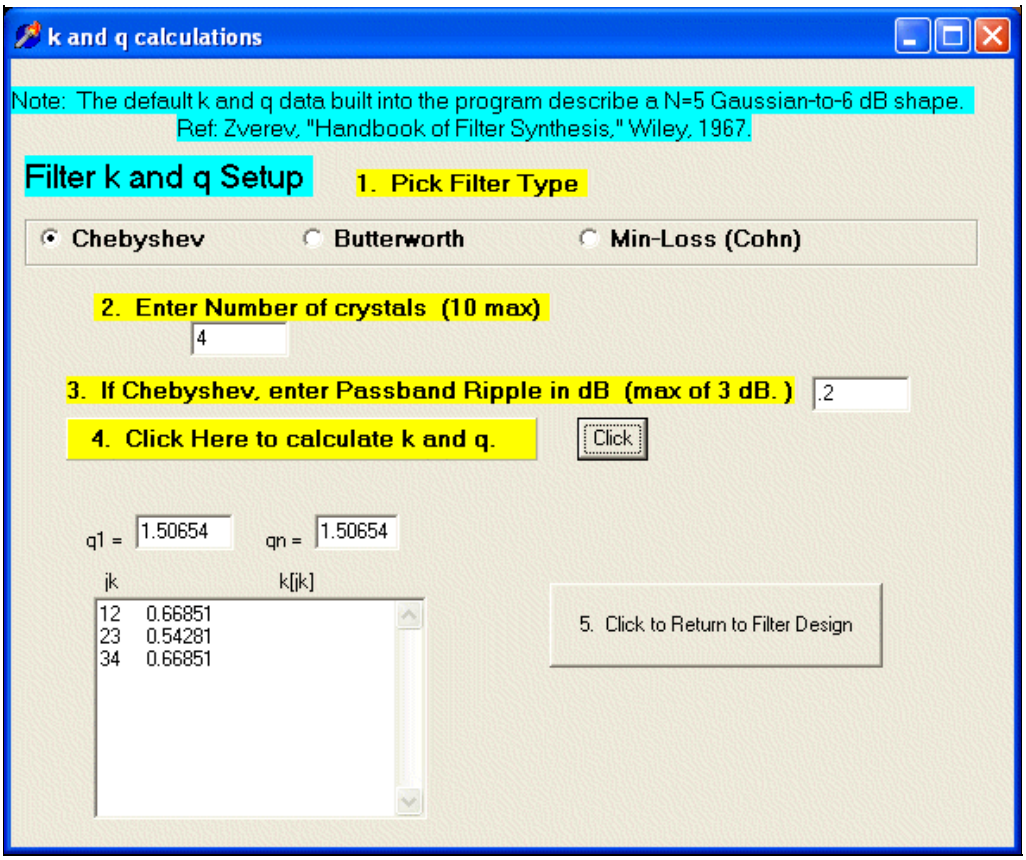
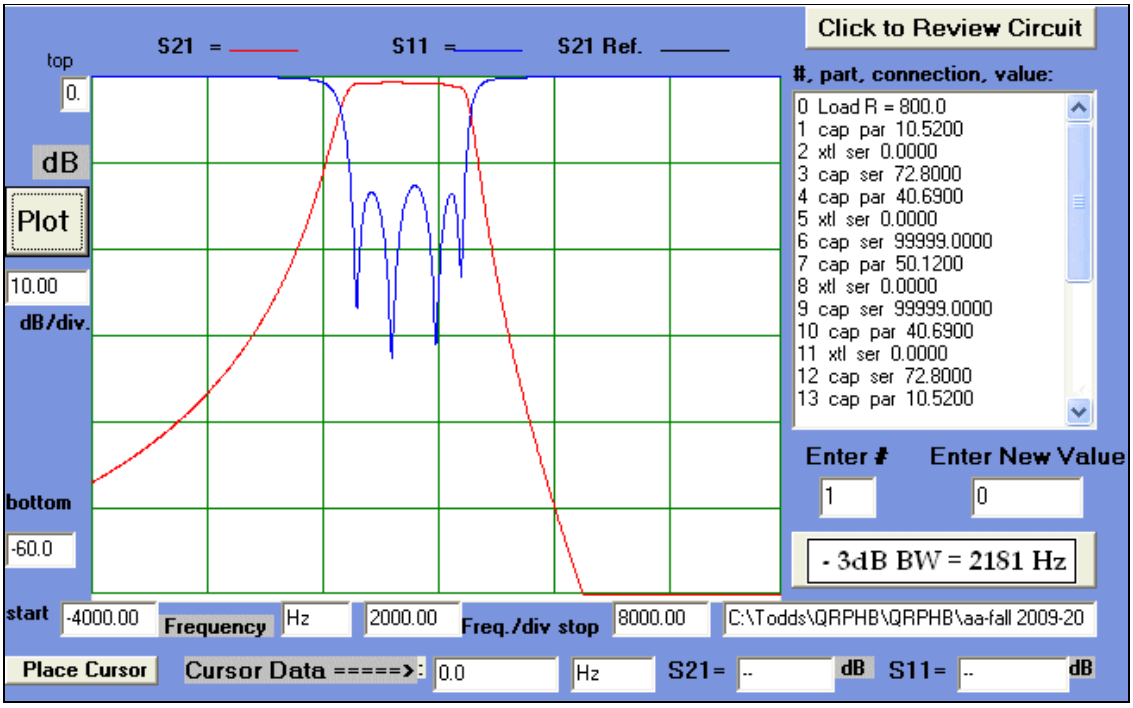
5. The Model 5 MHz SSB filter Design

Although this page is not about crystal filter design, an example follows for reference purposes.

For designing filters, the application xlad08.exe is a good choice. The following 3 screen captures show the raw design process and GPLA analysis of the model 5 MHz Lower Sideband filter shown earlier with C-par = 4.65 pF and Lm = .0578. There are some great articles in print and on the Web to study if you want to learn about filter design. The Ladpac software from EMRFD is excellent. My special thanks to Wes, W7ZOI for answering my questions about his software. From this information, I was able to make this web page.







6. Conclusion

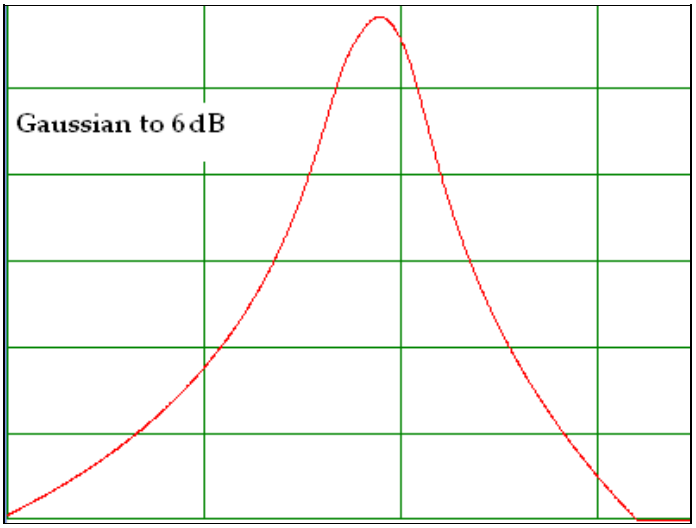
This web page presents a brief method to calculate crystal motional parameters and as required, to adjust crystal filter circuits to function optimally. This approach like my Java-script applet are simplistic to avoid the fear factor associated with crystal ladder design.

Listening tests are also valuable for assessing crystal filter function. Is the bandwidth as you expected? Does the filter ring excessively? Does it sound tinny? In the recent past, the crystal filters in 2 kit receivers/transceivers were tweaked as a favor to friends. *Please note, I have total respect for people who sell kits and appreciate the contribution they make to our hobby.*

The crystals of these Cohn type filters were removed and analyzed and the bandwidth was not as specified. In 1 case, the filter shape looked terrible in GPLA. Clearly, the kit sellers provided crystals which had markedly different parameters from those used by the original circuit designer. The I.F. filter capacitors were replaced with appropriate values and the R values were adjusted via either resistors and/or transformer

ratios and the improved filters sounded pleasant.

It is a real treat to listen to a receiver with a well designed crystal filter — sadly, I don't enjoy this experience that often.



In my opinion, the best sounding CW crystal filter design is the Gaussian to 6 dB. Some operators would never use such a filter in a contest-grade receiver as the filter skirts are not steep enough for them. There are tricks to make the stop band better (more like a Chebyshev response, but without the ringing), however, this topic is out of scope.

I sincerely ask for your feedback on the G. JavaScript Applet. Does it work correctly? How could it be improved? Can you contribute better code? Thanks and 73.

QRP — Posdata for January 2012 — More on Crystal Ladder Design

Important to both superheterodyne receivers and single sideband transmitters, crystal ladder filter design lies juxtaposed as both a favorite and feared RF design topic. Newer builders may lack math skills, and/or become paralyzed by the terminology — or lack the ambition to learn or apply good bench practices. Even a cursory Internet search returns many fabulous files to read — witnessing a crystal ladder filter design article explosion.

The difficulty characterizing and building filters has progressively declined since the advent of the first handheld computers — improvements in personal computers and filter design software allows astute builders to pursue even complex xtal filter response shapes in 2012.

In QST for July 1987. Wes, W7ZOI wrote **Designing and Building Simple Crystal Filters**. This article promoted Cohn or Min-loss filters and its intent was to transcend the math and measurement associated with xtal ladder filters of the day and allow builders to just frequency match some crystals, and then go experiment.

In my estimation, this article proved revolutionary — soon after, builders around the globe, Elecraft, and other kit companies embraced this technique/topology and the rest is history. *(I call it the paper that launched a thousand kits).* If you're a new builder and feel overwhelmed by the material on or referenced by this web page, please consider first obtaining this article.

Learning about crystal ladder filter design is time well spent.

In 2011/2012 I explored the works of 3 builders who share their work via the web and/or journals.

Horst Steder, DJ6EV and Jack Hardcastle, G3JIR

The Steder and Hardcastle works emphasize that we need to measure/calculate crystal L and C parameters, plus the coupling and tuning capacitances (not just frequency). Through emails with Horst, DJ6EV, I learned many things, but 3 stand out:

1. It's better to design a good filter than fix a bad one.
2. Careful measurement of your crystal parameters plus software design and simulation = the best chance for getting your desired performance.
3. Deriving motional parameters from a 3 dB bandwidth measurement remains a great way to characterize multiple xtals. *In my opinion, the G3UUR method is the easiest way to characterize a small batch of xtals.*

Some of the earliest references to modern crystal ladder design I've found were written by Jack Hardcastle and published in RadCom and QST — I later confirmed this by reading work by Wes, W7ZOI and others. Hardcastle's and Haywards' work proved foundational for the experimenters that followed including David Gordon-Smith, Chris Trask, Jim Kortge and many others.

Steder and Hardcastle's combined experience assessing and/or documenting crystal ladder design spans decades.

Their QEX article **Crystal Ladder Filters for All** may be legally downloaded from the ARRL website here.

Program download URL: http://www.arrl.org/qexfiles (select 2010, "11x09_Steder_Hardcastle.zip")

The QEX article describes Steder's Microsoft Windows™ program, methods to derive motional parameters, plus cites many important references. The main program calculates practical lower-sideband crystal ladder filters based on the exact equations published by M. Dishal in 1965. Hardcastle transformed these equations into a computer useable form in 1983 and Steder incorporated these equations into a modern, easy-to-use and interactive application with nice graphing and table displays.

For simplicity, the program assumes lossless crystals, however, the calculated values can easily be transferred into another simulation program such as GPLA to add or refine parameters such as loss resistance.

The main "Dishal" application calculates filters with Butterworth, Chebychev and constant-k (Cohn) properties and the so-called "QER" filter type by G3UUR (a low ripple version of the Cohn filter). Further; sub-programs in the top-level menu calculate xtal parameters (by both the G3UUR and 3 dB method), plus L-C impedance matching and ladder termination networks. An extensive help file well explains the program.

Iacopo Giangrandi — Introduction to Crystal Ladder Filters

Link: http://www.giangrandi.ch/electronics/crystalfilters/xtalintro.html

Iacopo (Jack) uses a transmission measurement to infer the motional parameters — inserting a series capacitance and measuring the series resonant frequency shift was also described in 1998 by Rolf-Dieter Mergner, DJ9FG.

His web article/applications provides what is likely needed by most builders — simple filter synthesis while avoiding expensive test gear. His filters plots/figures are spectrum analyzer measurements that I really like. Although his program can generate aymmetrical filters that some builders might not be used to, the frequency domain plots indicate proper function.
Giangrandi's filter design programs appear to be based on Jack Hardcastle's work and possibly content published in a paper by Patrick Magnin, F6HYE and Bernard Borcard, F3BB in Radio REF for April 1990.

I encourage you to try all the methods and applications mentioned to discover what works best for you. Don't lose heart, for characterizing crystals with a vector network analyzer is also a time-consuming endeavor and simple often = best on the QRP Workbench.

QRP — Posdata for August 2012 — Measuring Crystal Q

Prior to July 26, 2012, I could not measure Xtal Q. Why? I tried to measure crystal Q with the shunt-series tuned method (essentially the crystal acts as a trap and a step attenuator is used to calculate the insertion loss the xtal exhibits when centered in the notch) but failed because I could not precisely set the frequency with my homebrew L-C oscillator.

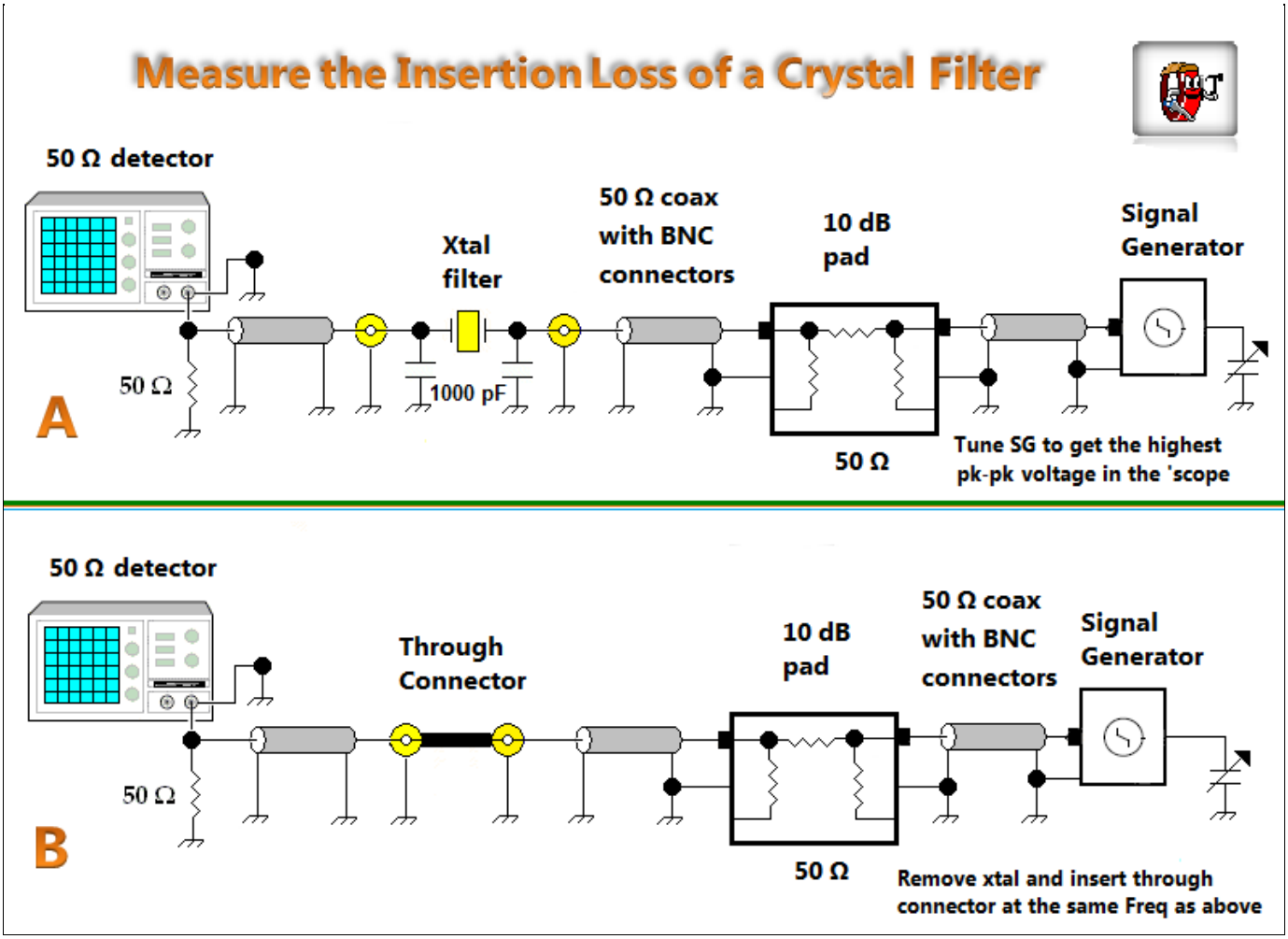
You really need a DDS or a Si570 based signal generator and preferably a spectrum analyzer to exact the measurement with the "trap method". The DDS is critical, while the SA only preferable — a power meter, or a 50 ohm terminated 'scope can work as the detector if a low-pass filter is placed just after the xtal.

On July 26, 2012 Wes, W7ZOI developed a simplified method and wrote a pdf file called **Simplified Measurement of Crystal Q** after feedback from myself and John Larkin about Q measurement difficulties without a digital-based signal generator. Unfortunately, this pdf file is no longer offered for download by W7ZOI. His method works well and I'm glad that as of now, I can *completely* characterize any crystal I own.

I present an experiment showing how I measured the Q of a 10 MHz crystal applying the new method developed by Wes, W7ZOI.

The crystal is evaluated as a N=1 low-pass filter resonated by a shunt capacitor at each end. I stuck with Wes' suggestion to try 1000 pF at 9 to 10 MHz. For lower frequency xtals he recommended trying larger value shunt capacitors. Just experiment with the shunt capacitance — if you use too high a C for a given crystal, your xtal low-pass filter bandwidth might get too narrow to measure with an L-C based signal generator.

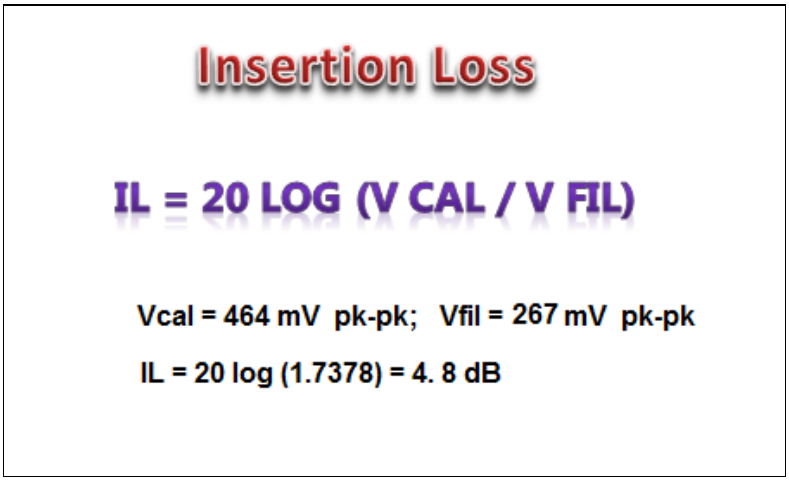
The following diagrams employ 2 programs from the Ladpac software that ship with EMRFD



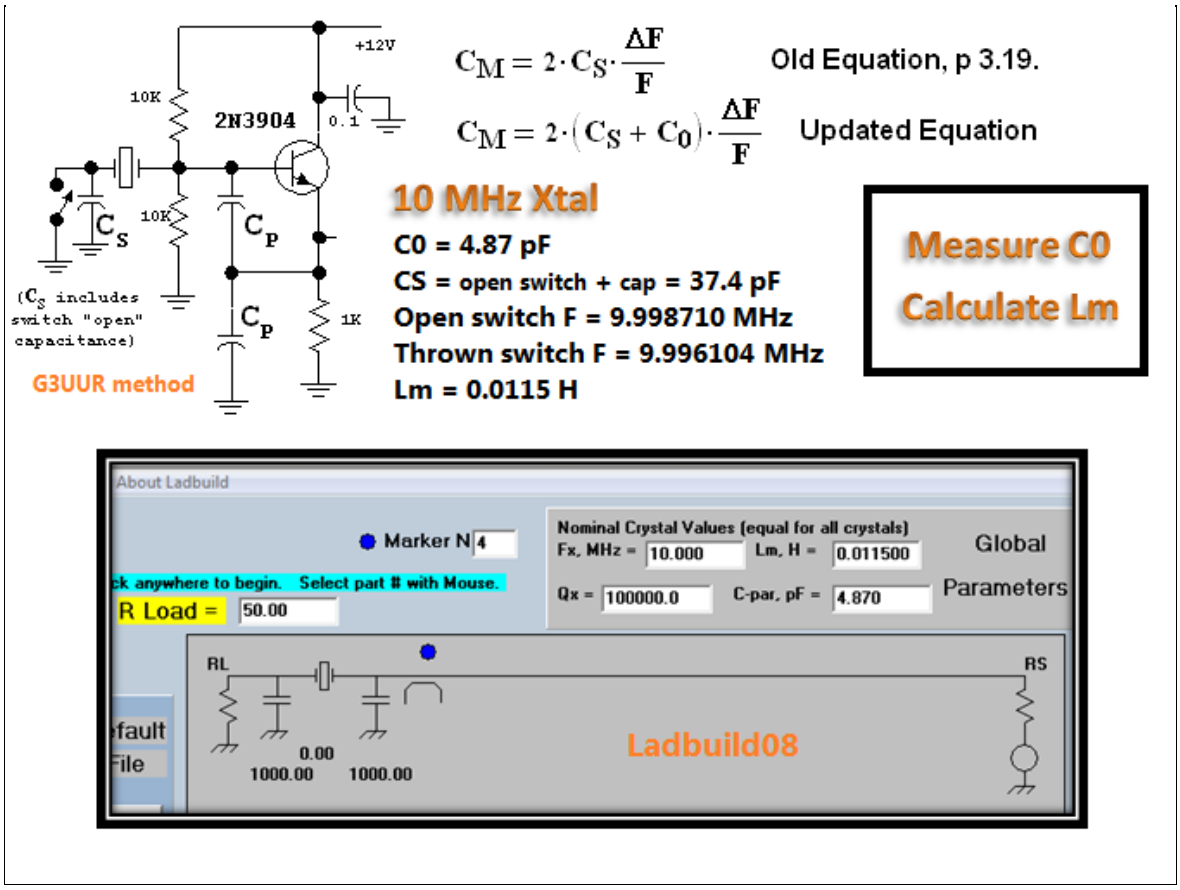
Above — Measurement of crystal filter insertion loss. In Part A, I carefully tuned my signal generator to get the highest peak-peak voltage in my 50 Ω terminated oscilloscope. I recorded this AC voltage as V Fil. In Part B, I removed the crystal filter board and replaced it with a BNC clad RF through-connector. I recorded this AC voltage as V Cal.

I discuss this standard method to measure the insertion loss or gain of a *device under test* in RF Workbench 2.

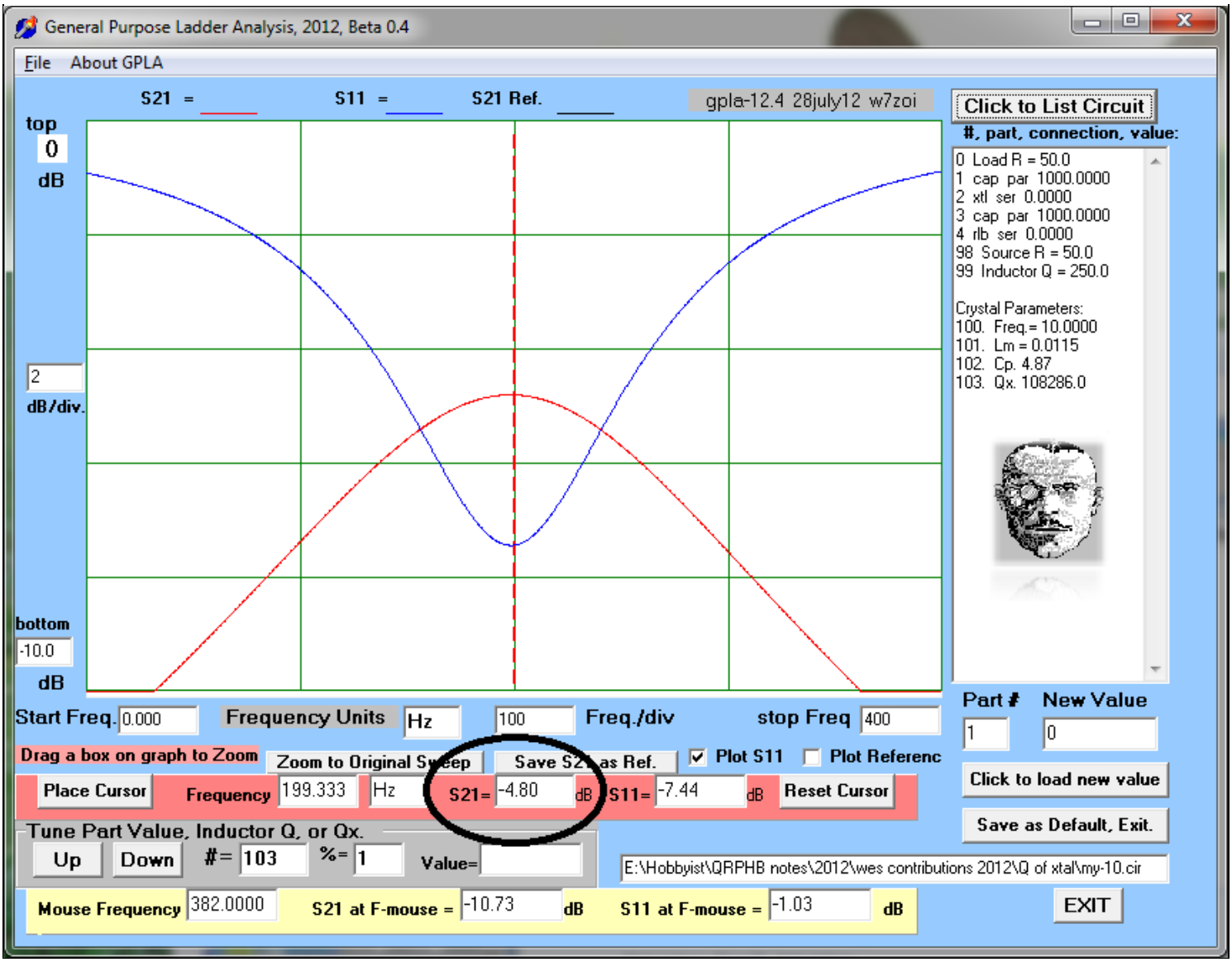
Even with the simplified method, you'll need a signal generator with good tuning resolution. My generator is shown on VFO 2011 as the 2.8-10.5 MHz Signal Generator. This is my version of the EMRFD Figure 7.27 generator.



Above — The formula for insertion loss using peak-peak voltages. With my 10.0 MHz crystal, V Cal = 464 mV pk-pk and V Fil = 267 mV pk-pk. The insertion loss of my crystal = 4.8 dB.

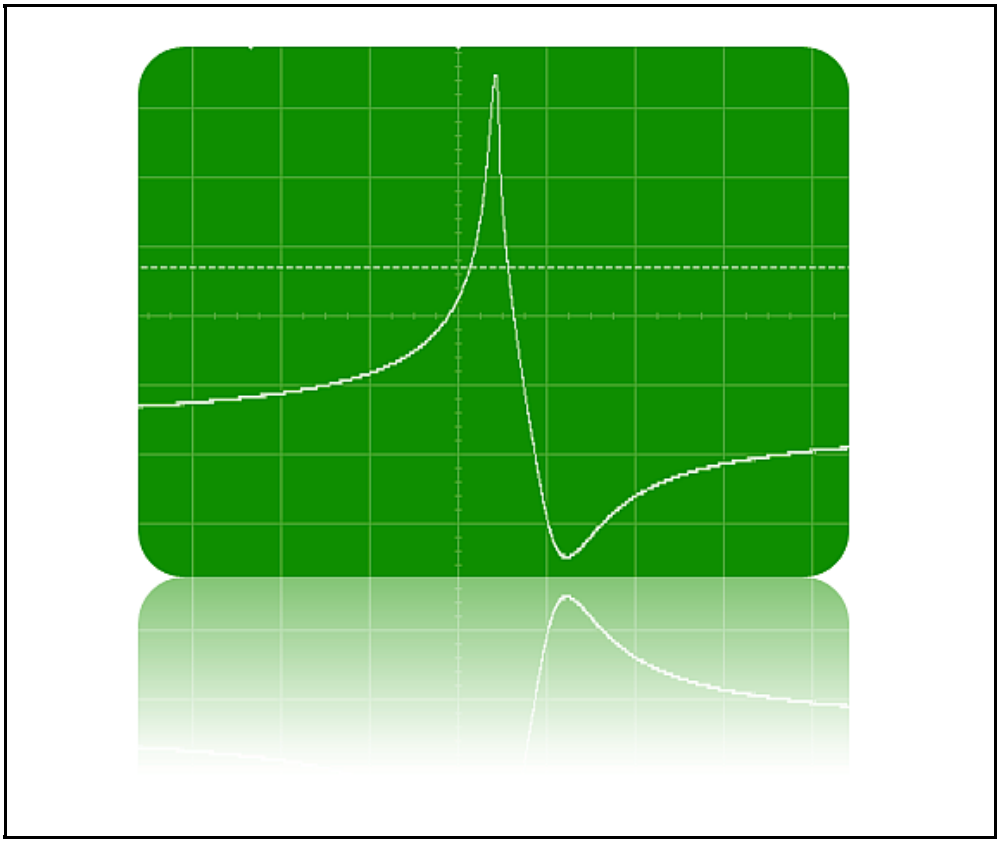


Above — First I measured C0 (C-par) and then with the G3UUR method, calculated the motional inductance of my xtal. Finally, I entered all the needed variables into Ladbuild08 to make a filter to analyze in GPLA.



Above — With GPLA, I adjusted the value for *Qx* up or down until my *centered* S21 value indicated -4.8 dB (the insertion loss of my crystal determined earlier). My crystal Q = Qx = 108286.

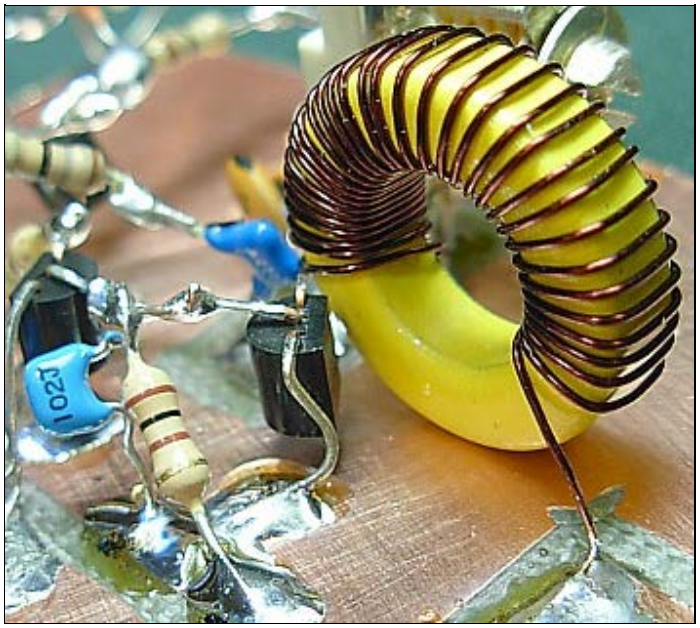
Above — 10 MHz Crystal filter breadboard.



Above — A sweep of the 10 MHz Crystal filter used to determine the Crystal Q

Amateur Radio Electronic Design

Electronic Hobbyist Circuits 2010



This page houses a collection of brief hobbyist experiments.

1. LM380 Power Examination

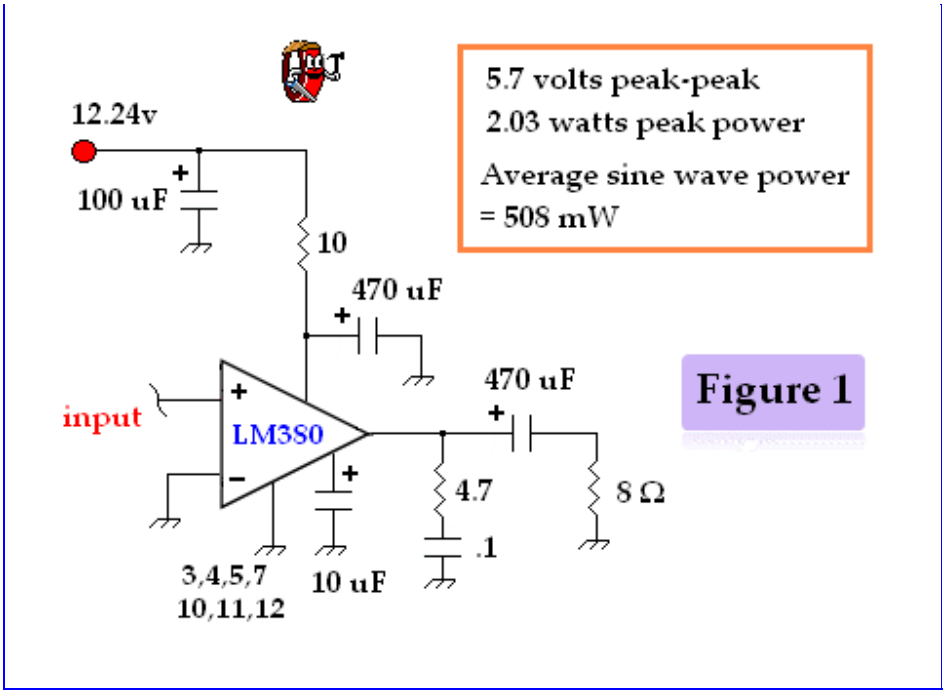
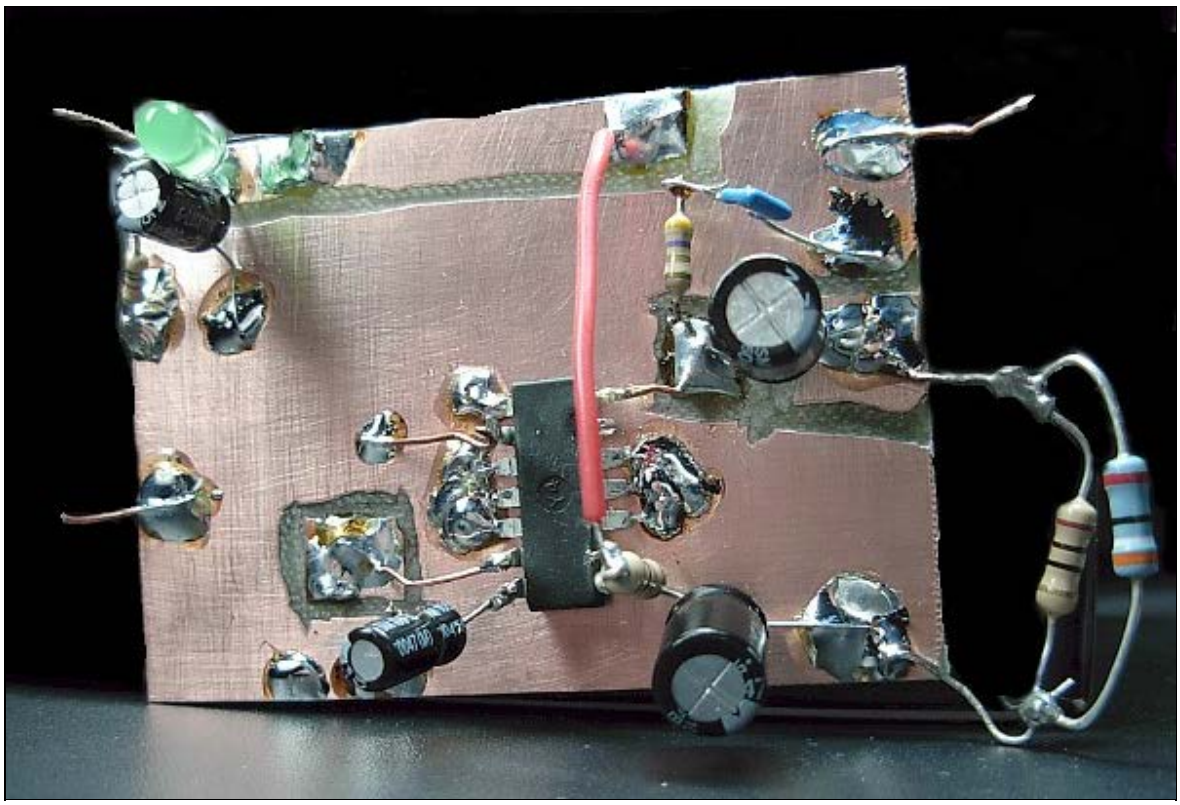
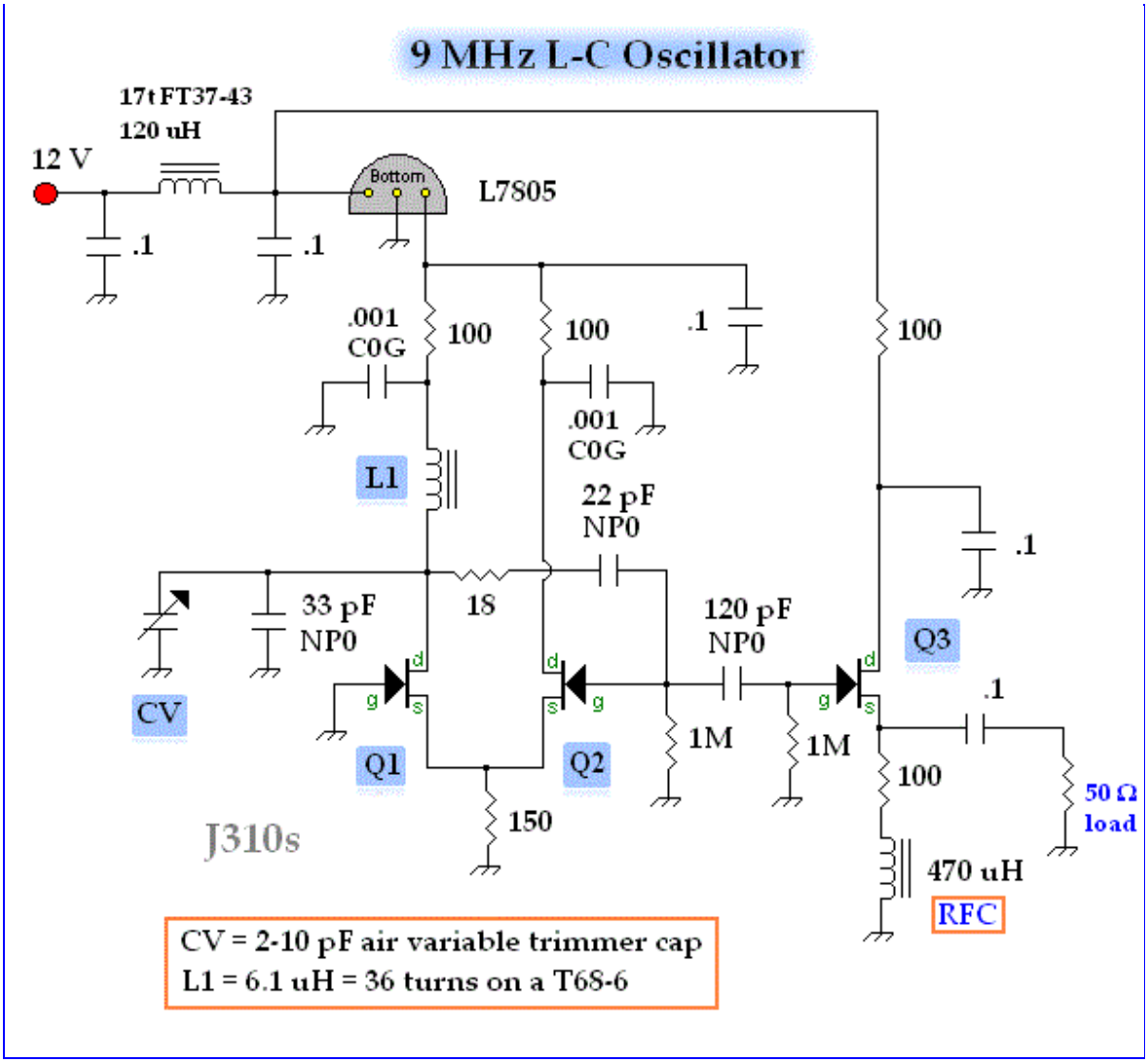


Figure 1 shows the test set up. This is a good part with an input impedance of 150kΩ. The gain is internally fixed at 34 dB. The average clean power was 508 mW. The test input frequency was 1018 Hertz.



The breadboard of Figure 1 is shown above.

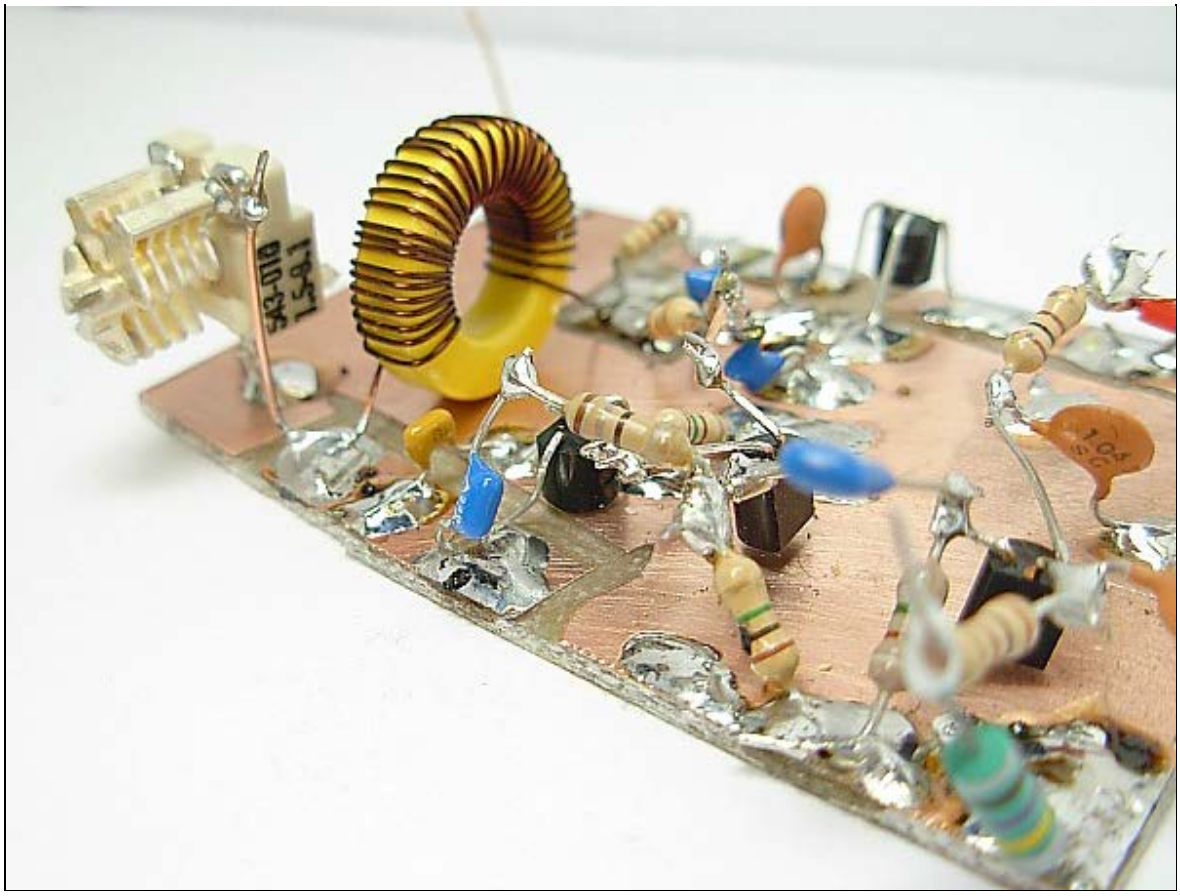
2. Wide Range L- C Oscillator



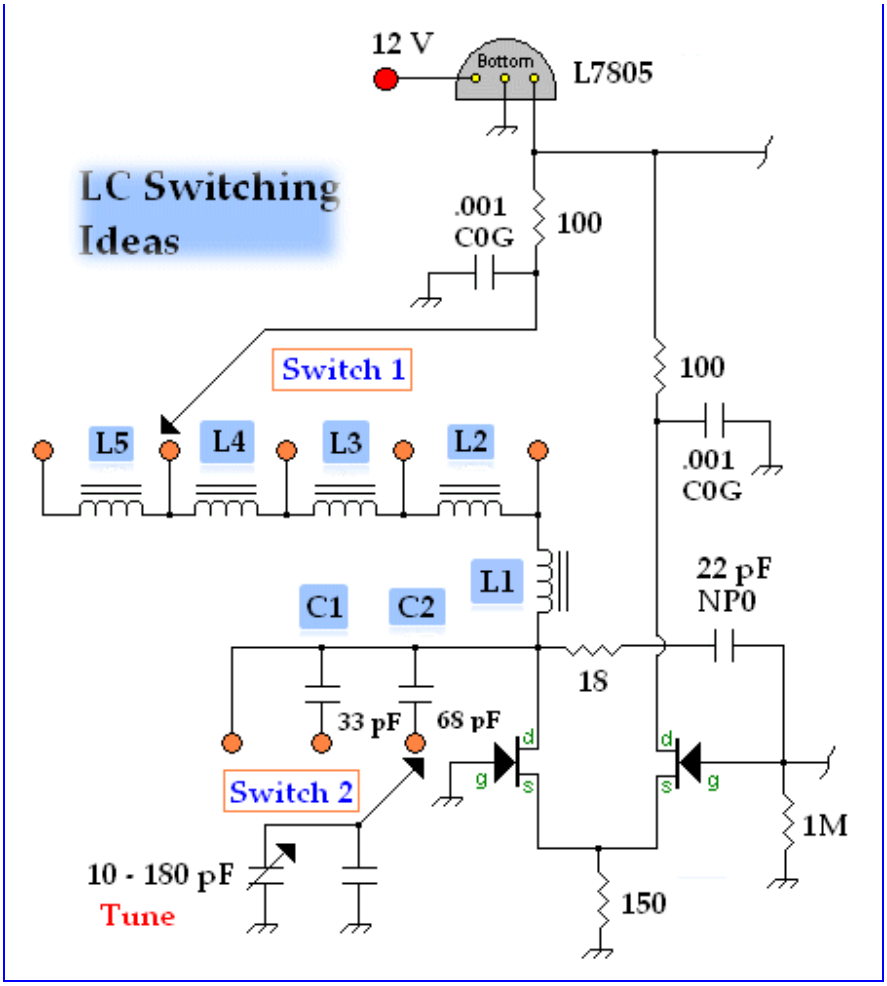
Shown above is a single frequency version of a VFO topology which allows a wide frequency range when additional switched inductors and/or capacitors plus a tuning variable capacitor are used. One good usage example would be a to use such a VFO to drive a bridge to make a wide range antenna analyzer. Q1 is essentially a common gate amplifier. The source is driven and the output is taken off the drain. This FET exhibits no signal phase shift. Q2 is a source follower that is AC coupled through that 22 pF capacitor The 18 ohm resistor is used to kill UHF parasitic oscillations. The Q2 follower also has no phase shift. Connecting the output of Q2 back to stage Q1 gives zero phase shift. The L-C tank will select the frequency where 0 phase shift is obtained. The tank will show phases other than 0 away from its resonance.

Q3 is an AC-coupled source follower to further buffer the VFO from its load. The RFC can be anything from your junk box, although it should likely be low Q. The low-pass decoupling filter on the the 12 volt supply path can also be anything reasonable. I wound mine using 17 turns on an FT37- 43 ferrite toroid. Its purpose is to keep RF from traveling down the 12 volts DC voltage wire to other parts of your circuit.

Any component connected to the L-C tank (at the Q1 drain, or the cold end of L1) can affect VFO tuning and drift. Temperature compensation will be necessary to achieve perfect stability. I use NP0 and C0G caps interchangeably. In the design shown, stability was good and the output had low measured distortion. This VFO will pretty much oscillate with any reasonable L and C values in the tank circuit. I found frequency stability was a little better with a higher L to C ratio. This is a great experimenter's circuit. One version built oscillated at 150 MHz.

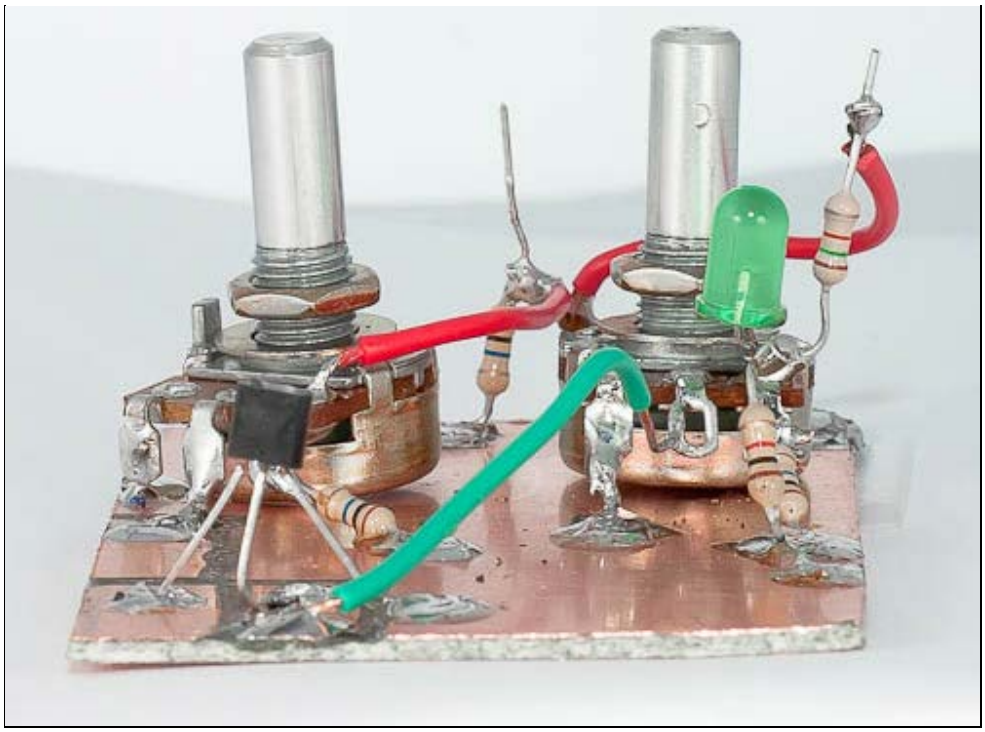


The breadboard of the above schematic. Pull the wire on your #6 powdered iron toroids tight to prevent air gaps between the toroid and the wire. Number 26 gauge wire was used on L1 as shown. High Q tank parts will garner the best results.

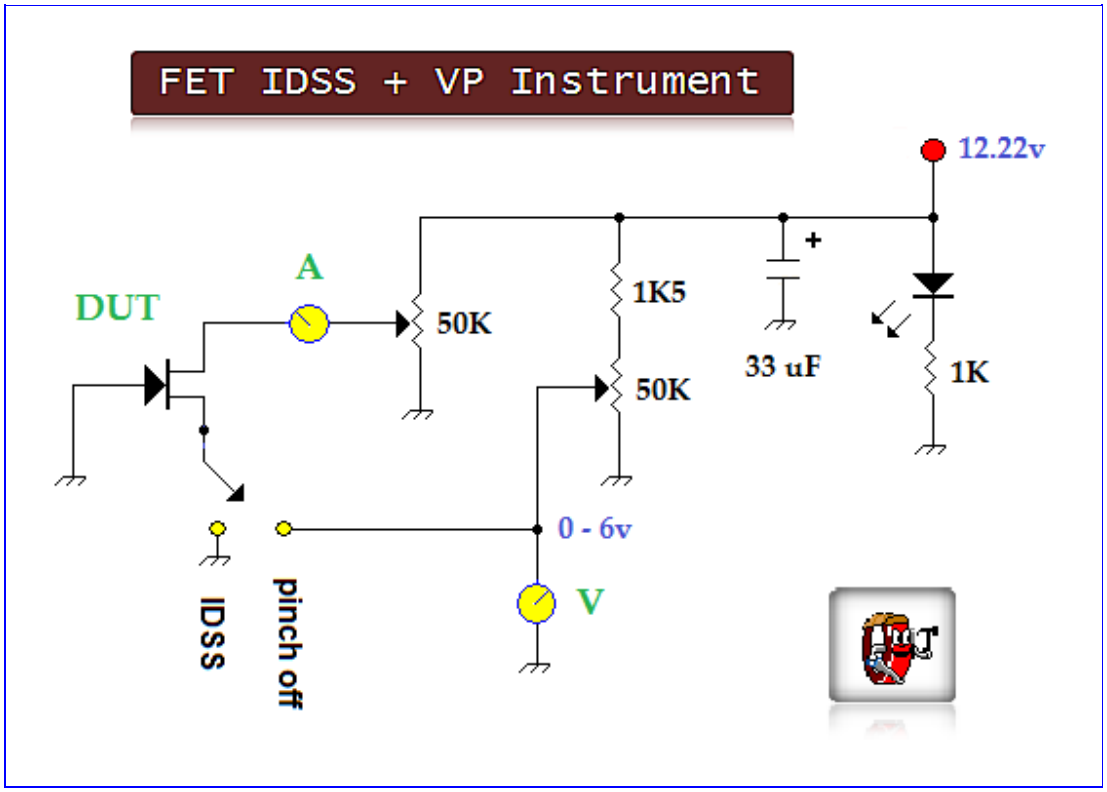


Some potential switching ideas are presented above. The builder is in total control of the tuning range and must calibrate the L and C values according to needs and the parts on hand. Output power will vary according to the L-C ratios and some designs include automatic signal amplitude leveling and/or RF gain controls.

3. FET Matching



I find matching high IDSS FETs like the J310 to be a pain. I generally matched them for IDSS and occasionally for IDSS and VP. Observations that when the IDSS of 2 or more FETs match, their pinch-off voltage (VP) also matches, led me to not measure VP. In addition, the variability of VP measurements causes me distress. Click here for a tutorial if you don't understand the terms IDSS and VP.



Above — The device I use to measure IDSS and VP. From Ken Kuhn's web site.

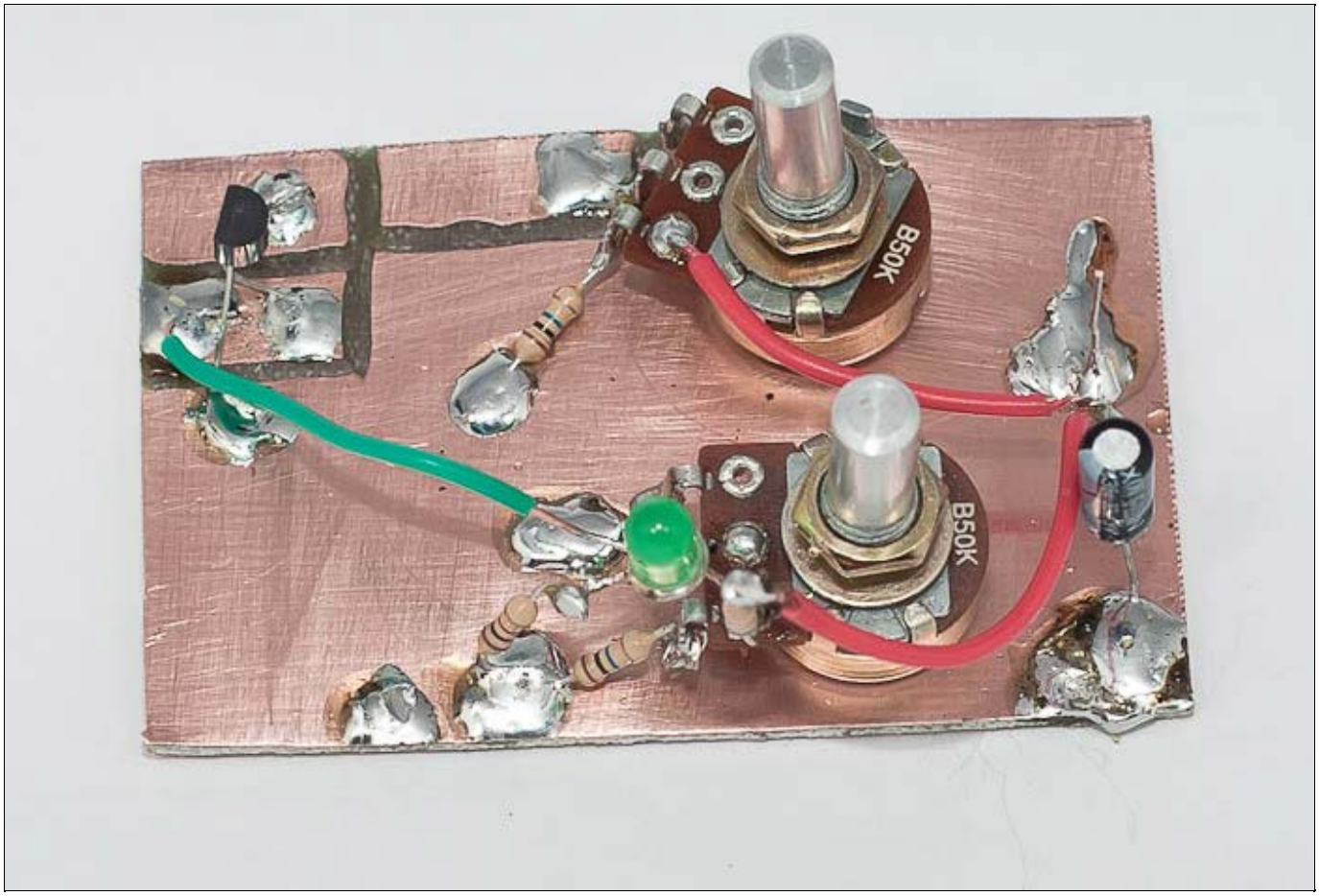
Conceptually IDSS and VP aren't difficult to understand — measuring them is another story. With the above device, first IDSS is measured; the final drain voltage potentiometer setting is left and then I measure VP. While measuring IDSS in high IDSS FETs, heating can occur and you may actually see current start to drop as you increase the drain voltage (negative temperature coefficient). On J310 specification sheets, the manufacturers state they pulse the current during measurement to prevent heating. While performing IDSS measurements, I am fearful of destroying the FET I am trying to characterize! Measuring VP is also problematic.

I have tried 3 methods to quantify VP:

Adjust the 0 - 6 volt supply until I think the current goes to 0. Serial measurements 1 day apart can vary by a variation of as much as 0.5 volts; it's quite subjective.

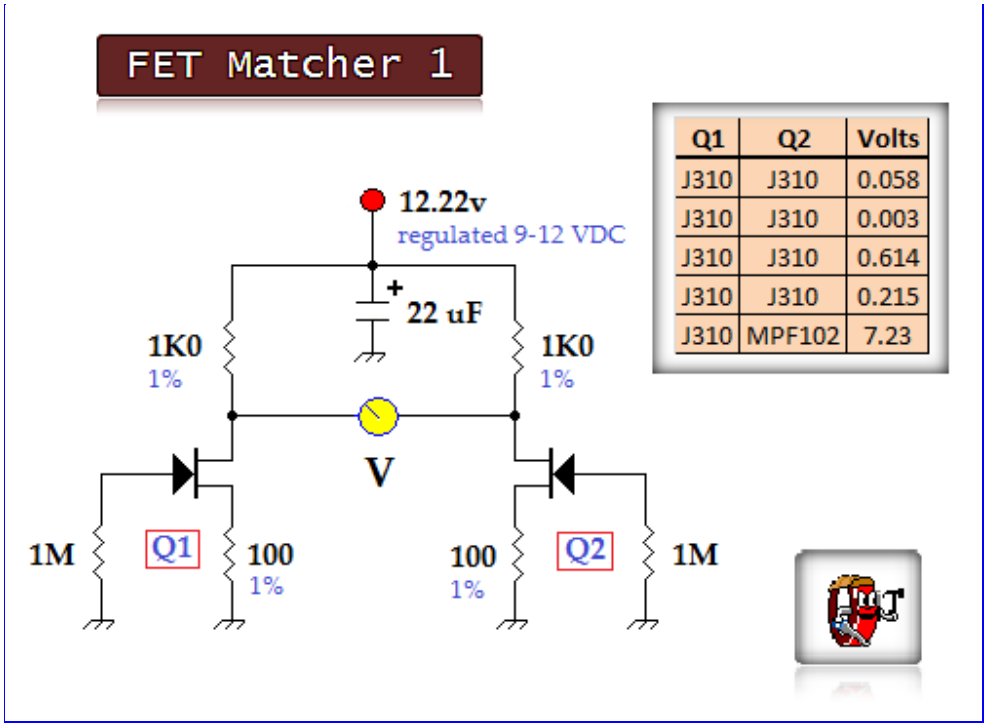
Adjust the 0 - 6 volt supply and measure the gate voltage that produces a drain current somewhere between 0.1 and 1 percent of IDSS and declare that to be VP.

Adjust the 0 - 6 volt supply so the ammeter reads ¼ IDSS and multiply this voltage by - 2.0. Refer to Ken Kuhn's site for details. Although reasonably accurate, the second order math is only a rough approximation — the real math is impossible to do by hand as it involves fractional exponents and these exponents and other factors vary as a function of the physical JFET geometry.



Above — The breadboard of the device I use to measure IDSS and VP. There is no actual switch, I either ground the green wire to the copper board, or tack solder it to the 0 - 6 volt potentiometer wiper. 10 megohm resistors plus the pot ground wire anchor each pot to the copper clad board.

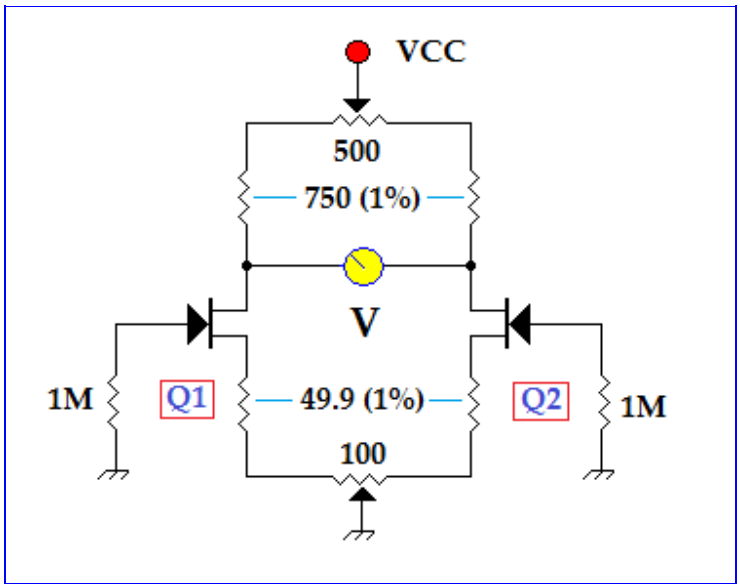
All 3 methods to quantify VP frustrate me. There must be a way to match J310s or other FETs without characterizing them. I frequently collaborate with readers to problem solve and learn. A potential solution contributed by a supportive reader follows:



Q1	Q2	Volts
J310	J310	0.058
J310	J310	0.003
J310	J310	0.614
J310	J310	0.215
J310	MPF102	7.23

Above — A bridge is used to match a pair of JFETS. It's often best to match devices in a circuit that closely resembles the one that you intend to use them in. The differential output of each drain is measured by placing a DVM lead on each drain and recording the voltage. Generally, I stick a FET in the Q2 slot and put FETs from my parts bin in the Q1 slot to match it. The results of 5 different FET pairs are tabled above. A match <= 50 mV is probably acceptable and in 1 case, I found a match of 3 mV! You can match 1 FET with many using this device.

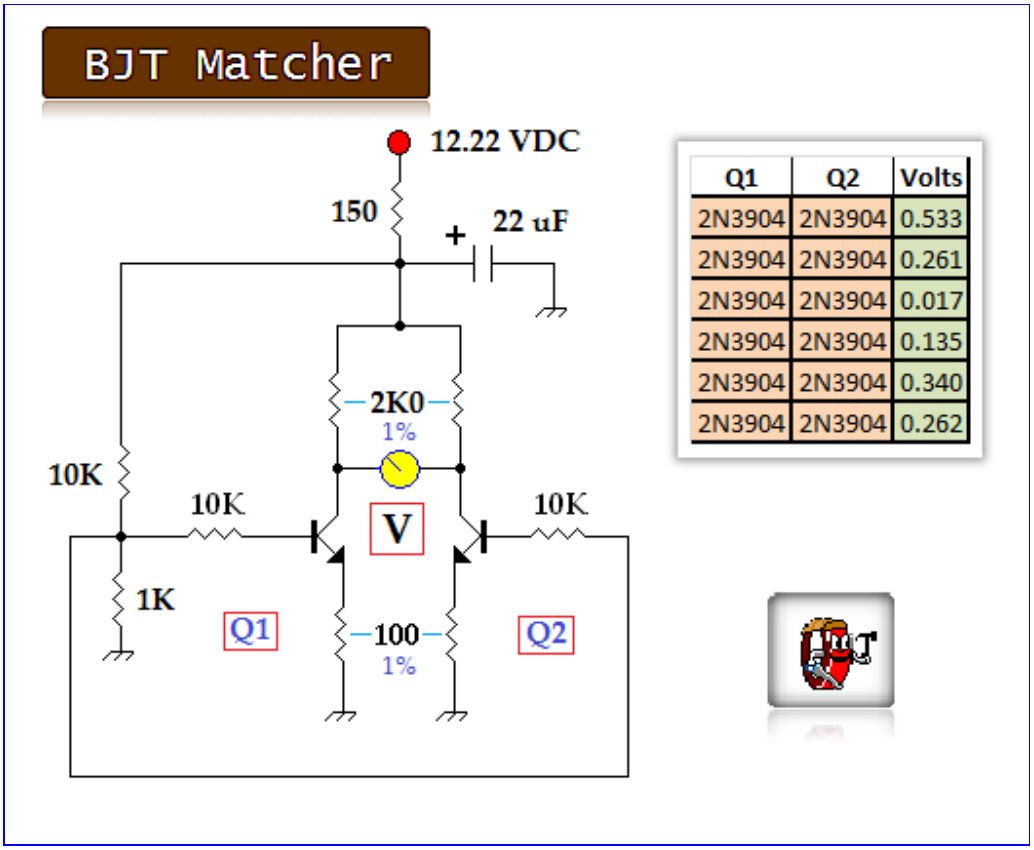
Note the poor match when an MPF102 and a J310 were tested. 1% tolerance resistors are recommended for the bridge.



Above — A set and forget precision bridge using trimmers to establish a perfect DC match on both halves. If you don't have 1% parts, the trimmer resistors offer a solution. You can place a trimmer at either the drain or source end as shown and just use 5% resistors. Calibrate each half of the bridge with your ohm meter. I cover bridges on this web page if you need more information on them.

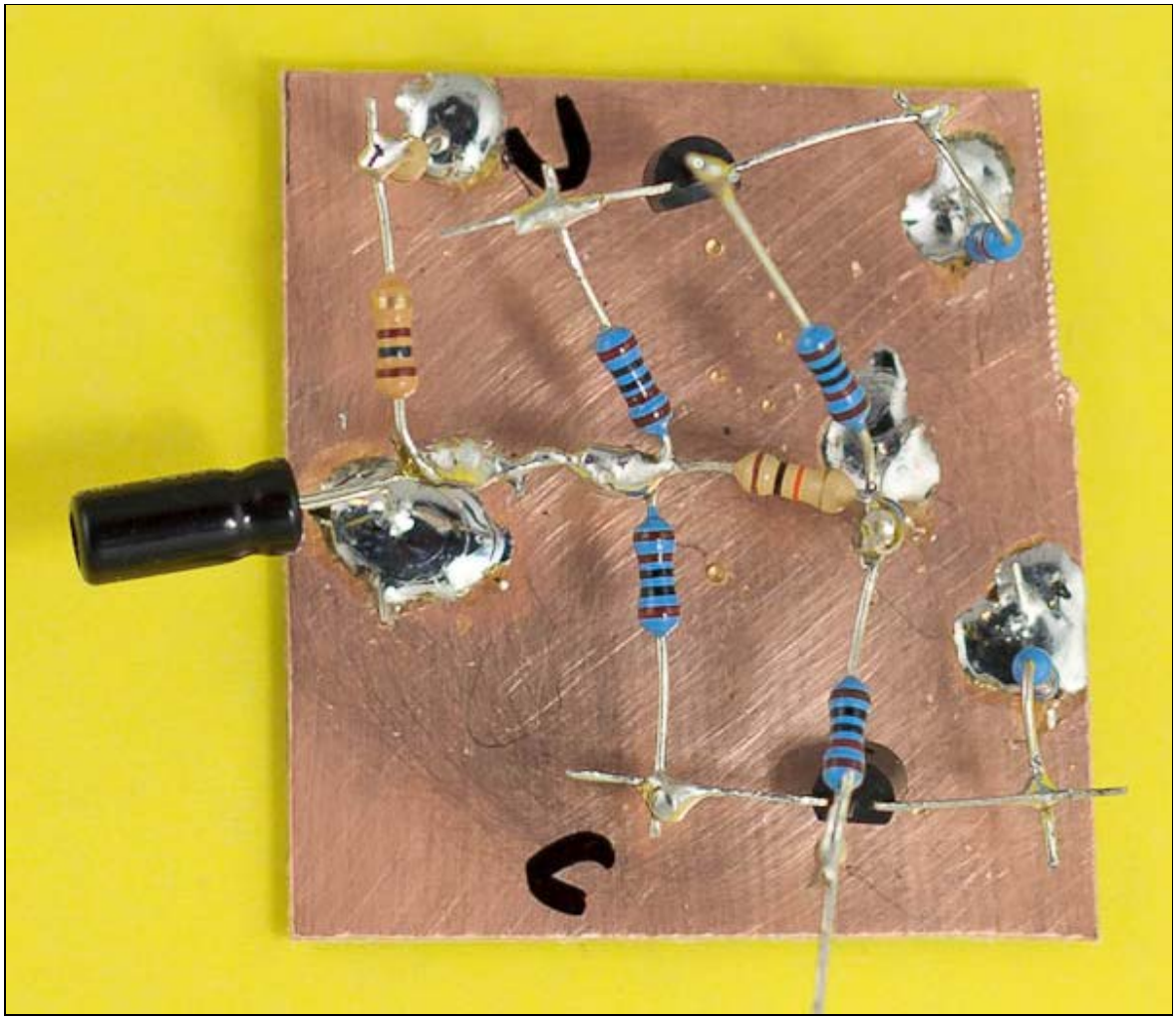
Above — A differential FET matcher breadboard. This version had a 9.1M gate resistor on Q1 by accident, although it made no difference to the experiment, as no DC gate current flows.

4. BJT and Diode Matching

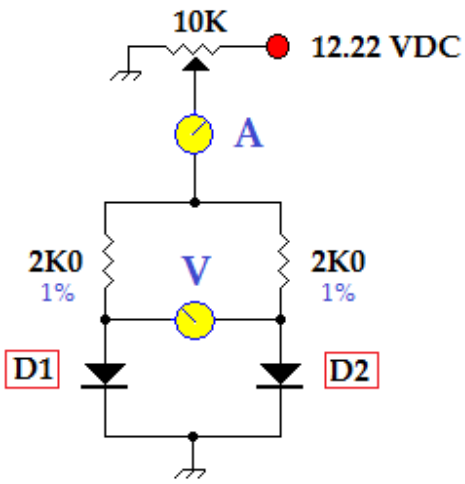


Q1	Q2	Volts
2N3904	2N3904	0.533
2N3904	2N3904	0.261
2N3904	2N3904	0.017
2N3904	2N3904	0.135
2N3904	2N3904	0.340
2N3904	2N3904	0.262

Above — Differential BFT matcher. Differential voltage matching works for bipolar junction transistors and diodes also. PN junctions are thermally sensitive — let them stabilize before testing. I measured 6 2N3904 transistors and the tabled results remind us why this transistor isn't the best choice for matched BJT circuits.



Above — The breadboard of the differential transistor matcher. Any reasonable resistor values will work fine, but don't run the current too high — my circuit has a nominal emitter current of 2.25 mA.

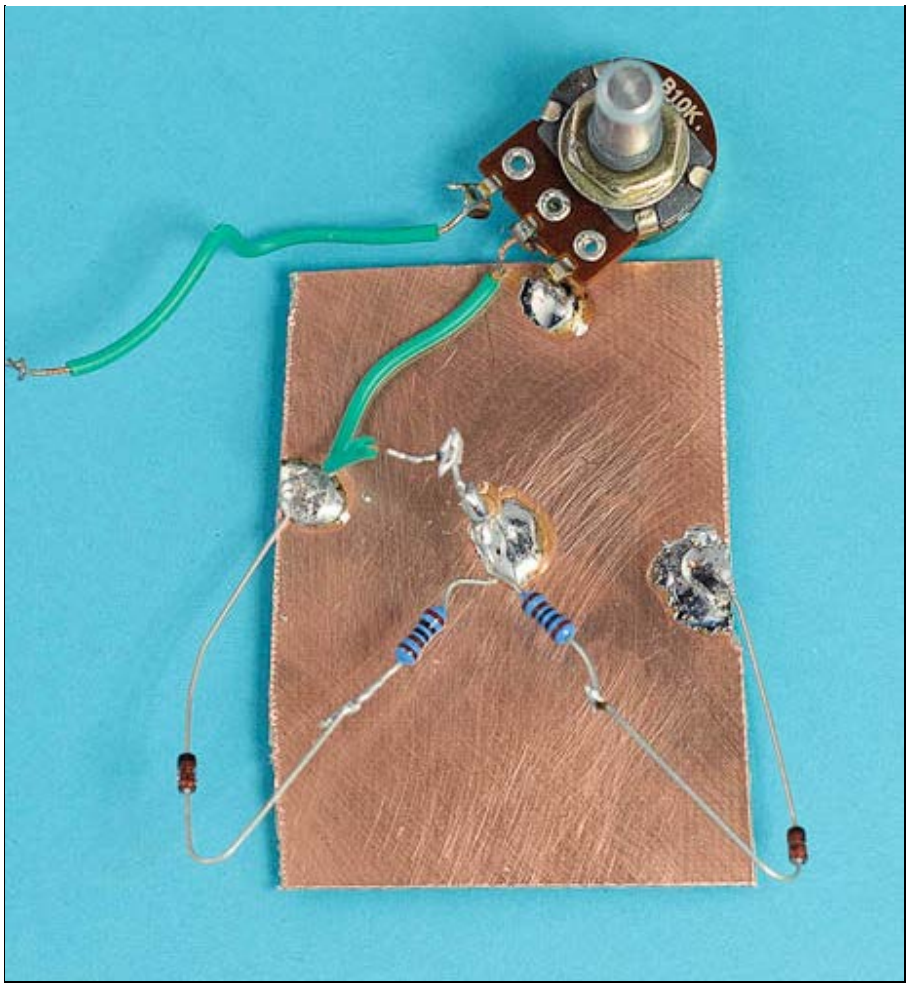


D1	D2	Millivolts
1N4148	1N4148	2.4
1N4148	1N4148	4.1
1N4148	1N4148	6.6
1N4148	1N4007	18.1
1N4148	Schottky	315
1N4148	Ge	169
1N4148	1N4148	6.8
1N4148	1N4148	4.9

Measurements with each diode current @ 77.3 µA

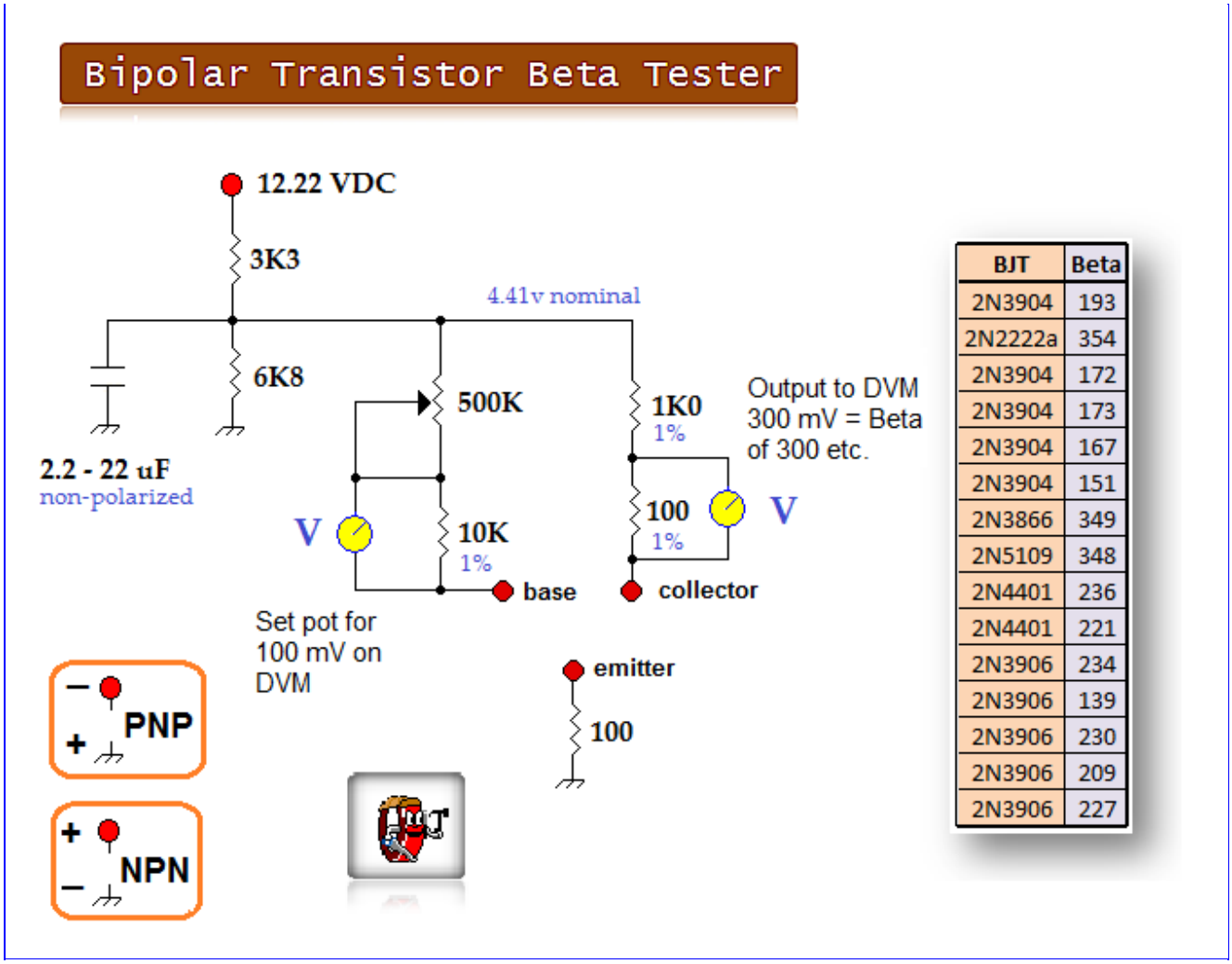
Above — The differential diode matcher. The 10K pot allows you to vary the current to suit your needs. The ammeter reads double the diode current. Considering only 1 diode; the current ranges from 0.01 to 5.82 mA. You can drop the 1% metal film 2K0 resistors down as low as 470 ohms or so if you need serious current. Tabled are some measurements performed with the 10K pot dialed to give 0.636v (77.3 uA per diode). This provided excellent sensitivity. Measurement was performed on my DVMs 200 mV scale. Builders might experiment with diode current to assess measurement sensitivity and linearity or even to match the diodes across a range of current values.

Clear glass passivated diodes can be affected by light — photons will pass through the glass and knock electrons through the barrier. Ensure each D.U.T is exposed to the same amount of ambient light. Some microwave detector diodes can be damaged with as little as 1 mA of current. Replace the 2K0 resistors with 10-15K resistors, or lower VCC to prevent damage .



Above — Breadboard of the differential diode matcher. The 2K0 resistors are suspended by a vertically mounted 10M stand-off resistor.

5. Bipolar Junction Transistor Beta Tester



BJT	Beta
2N3904	193
2N2222a	354
2N3904	172
2N3904	173
2N3904	167
2N3904	151
2N3866	349
2N5109	348
2N4401	236
2N4401	221
2N3906	234
2N3906	139
2N3906	230
2N3906	209
2N3906	227

Above — Disappointed with the transistor beta testers in our common, low-cost digital multimeters, we did the logical thing; designed and built our own. This collaborative project was more an experiment with BJTs than anything else. It's about as simple a beta measurement device as you can make and still get good results. Preventing damage to our parts inventory underpins this design — the 100 Ω emitter resistor plus ~ 10 microamps of base bias keeps the IC low to help avoid smoke since most new small signal transistors have a beta of 100-400.

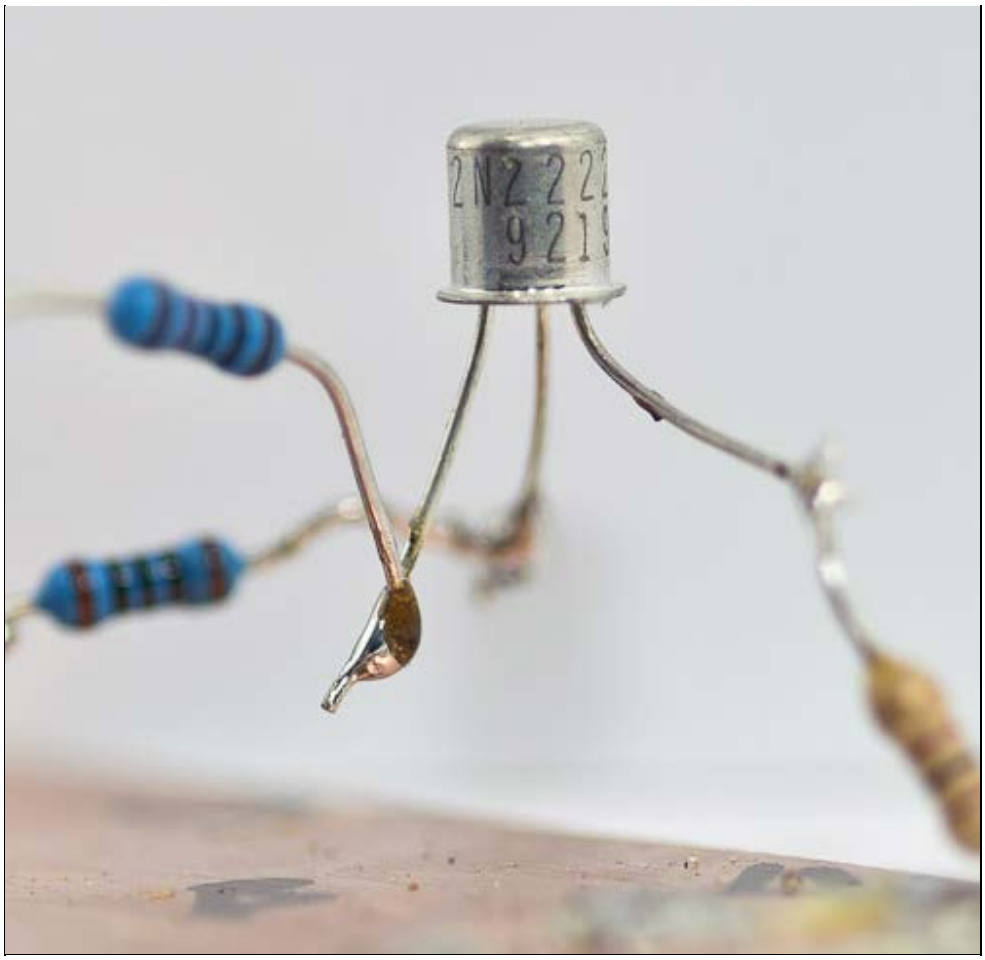
Ensure the correct polarity for PNP versus NPN transistors. The voltage divider targets 5 volts using a standard ~12 volt supply; I just used whatever resistors were handy and ended up with the 6K8 — 3K3 pair. VCC should be regulated. Perform the measurements with a single multimeter allowing time for stabilization.

To use: Set the potentiometer so that the voltage drop across the 10K resistor is 100 mV. Then move your DMM leads to the 100 Ω resistor and measure the beta. This device measures beta, the static gain at DC.

Measuring beta is a bit inexact since beta is affected by so many variables as follows:

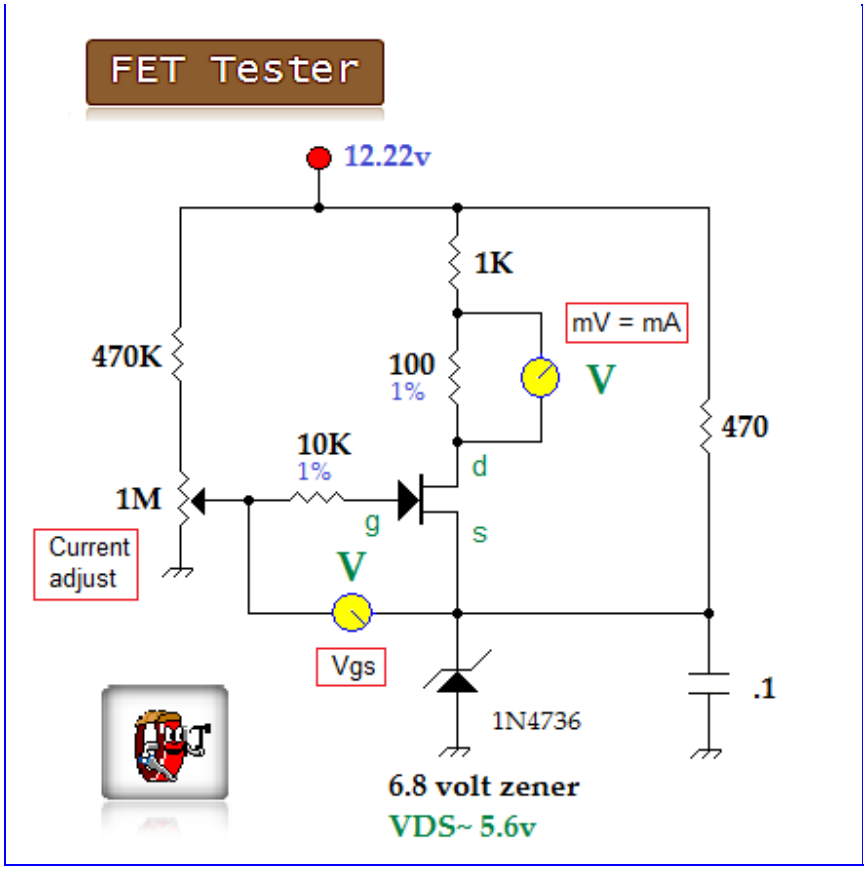
- Beta tends to be low at low operating currents and rises and plateaus for medium currents and then falls at higher currents.
- Beta tends to increase with temperature.
- Beta is affected by the voltage between the collector and emitter -- this is a weak effect except when the voltage is very small.
- The beta can vary as the battery depletes in DMM beta testers.

Above — Breadboard of the QRP beta tester. We hope you have as much fun with this circuit as we did.



Above — Testing a 2N2222a. We found measuring transistors to be very instructive; comparing our results to specification sheets, handling the different BJTs, gaining experience — all part of the exciting world of bench electronics. Such circuits, although simple, are great for both learning and design.

6. FET Tester - VGS Measurement Tool

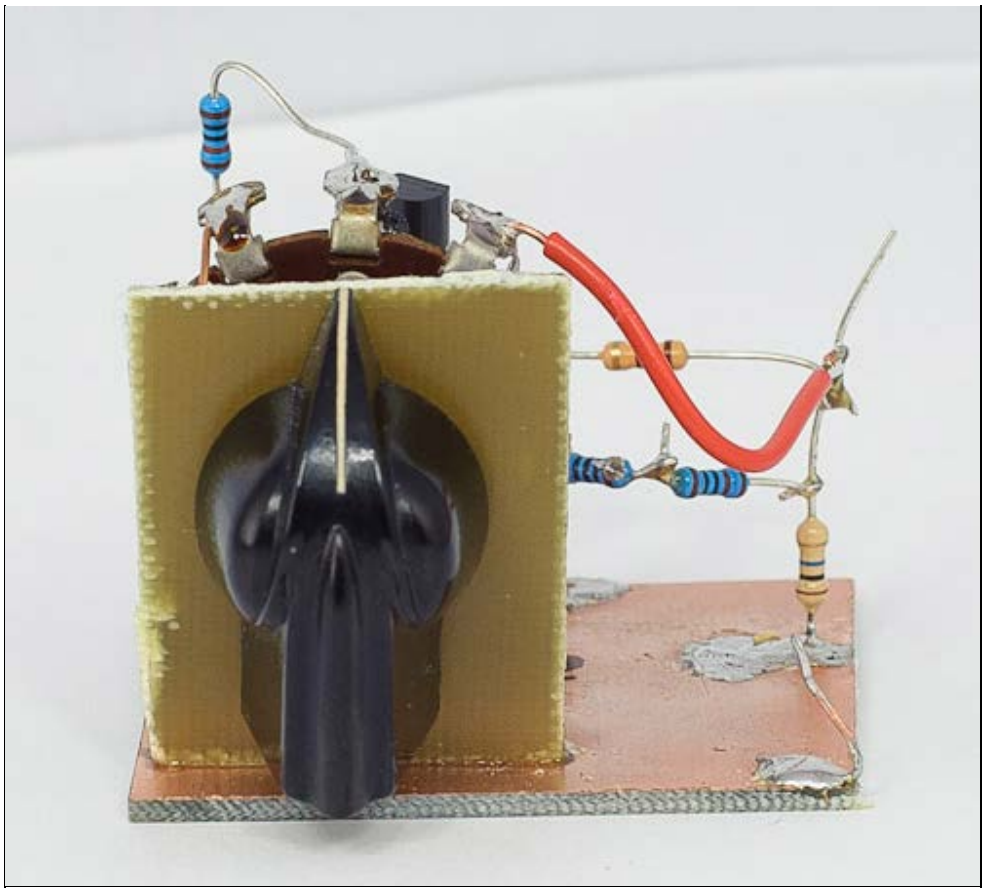


Above — A FET VGS Tester. The final collaborative output from late December 2010: a device to examine VGS. A zener diode value from 6 -7.5 volts or so should work okay. VDS should be greater than the nominal VP for greatest accuracy. The VDS using 7.5 volt zener diode was 4.85 volts. The range of VGS on my breadboard was roughly 6 to -6 VDC without the 470K range limiting resistor; it's up to you if you want to limit the VGS range.

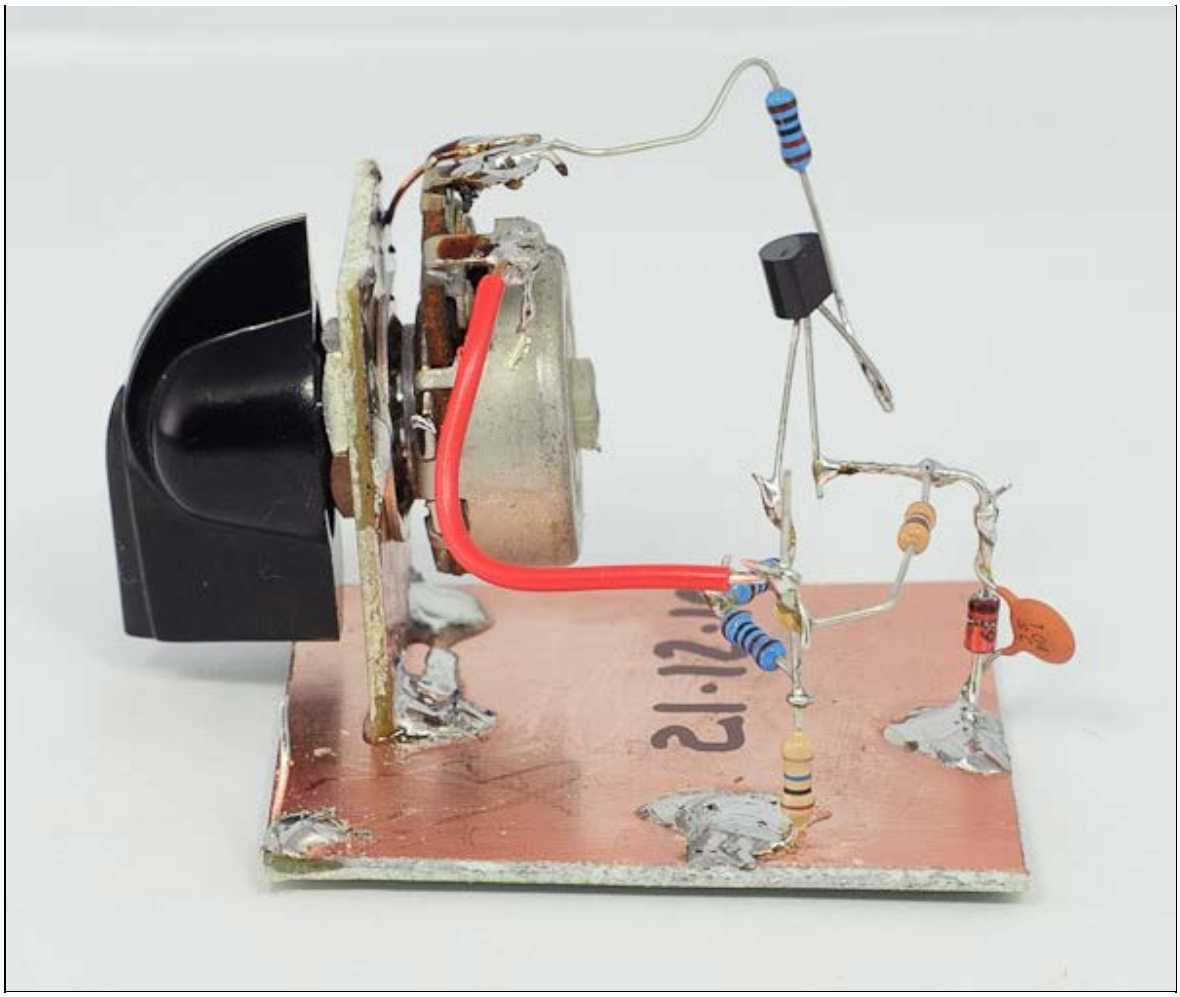
Using the mV scale on your DVM, potentiometer adjust the current until the meter just reads 0. Then move your leads to the VGS test points to read the pinch-off voltage.

The 0 current point will only ever be close. The interesting issue with the 0 point as defined is it never goes to true zero — just zero enough — that point may be hidden if your meter scale truncates lower current measurements to zero. If you're able to repeatedly get consistent VP measurements with whatever method your using, then it's likely accurate.

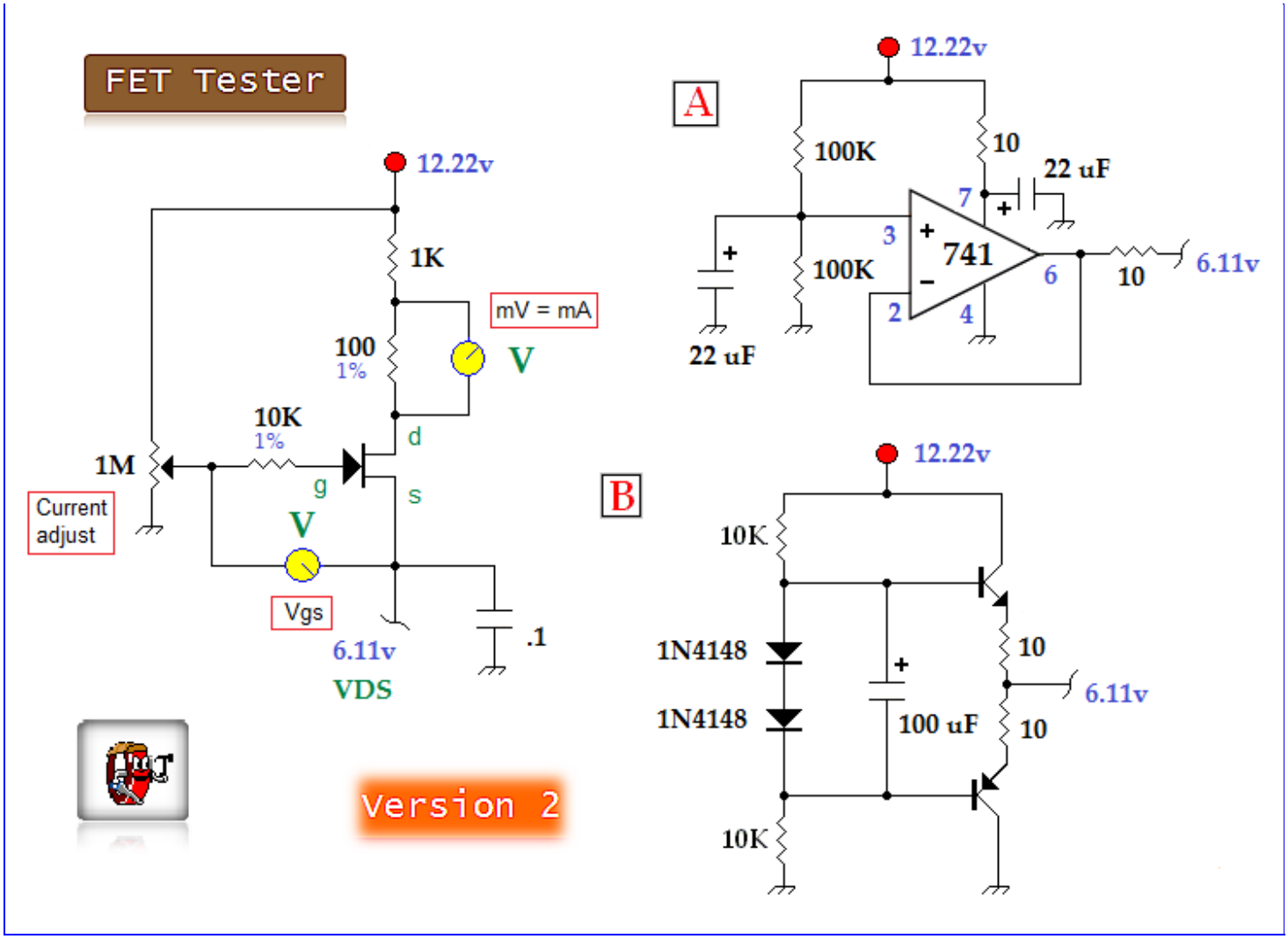
We thought about measuring gate leakage current across the 10K gate resistor, but accurate measurement is impossible with this device. The gate leakage current is going to be in the low nanoampere region at the very highest unless the JFET is bad. The expected voltage across the 10K resistor would be a few 10s of microvolts. A better way to measure gate leakage is to use the 10M resistance of a DVM instead of this resistor. The DVM then acts as a current meter by measuring the voltage across the 10M resistance. A reading of 200 mV would mean a current of 20 nA.



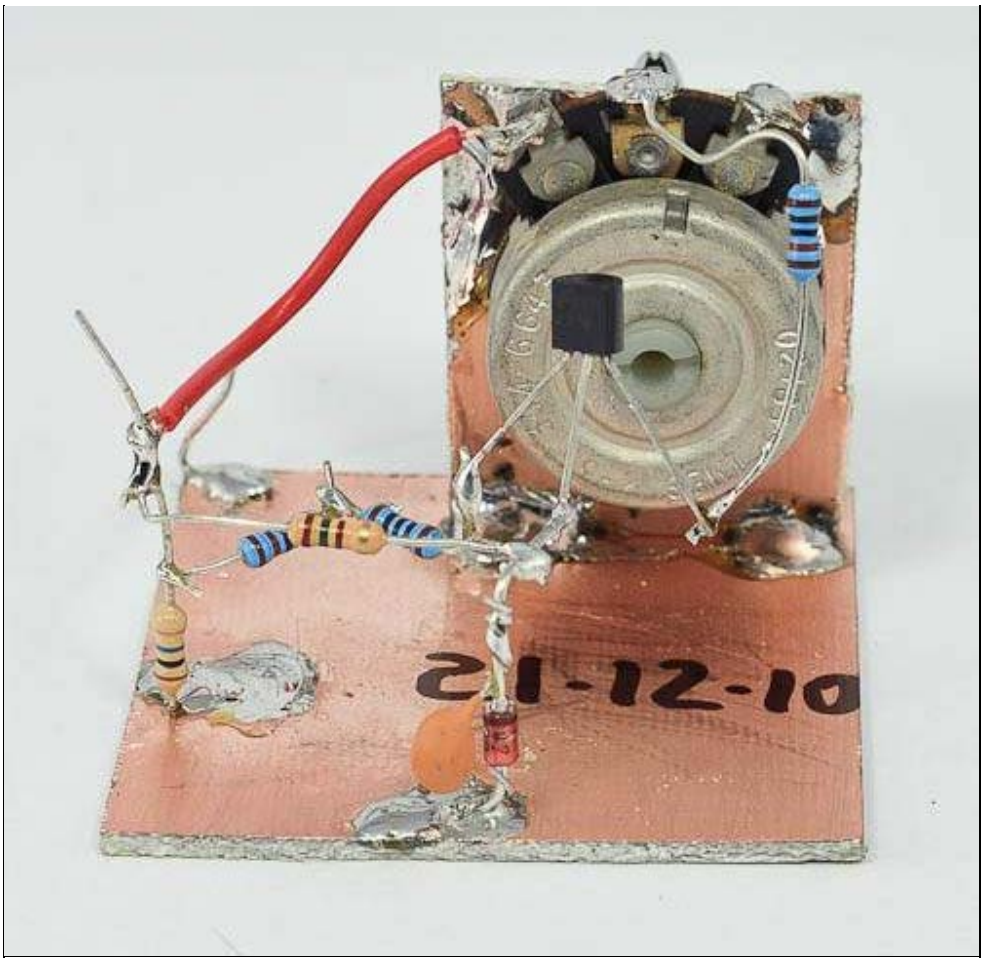
Above — Front view of the FET tester with a "chicken head" knob on the potentiometer.



Above — Measuring a J310 with the FET tester prototype breadboard.



Above — Instead of a zener diode, low impedance VCC/2 splitters are shown as A and B; the zener diode is easier.



Above — Reverse view. I didn't bother with the 470K VGS range limiting resistor in my breadboard.

RF — Test and Measurement

RF Workbench Page 1

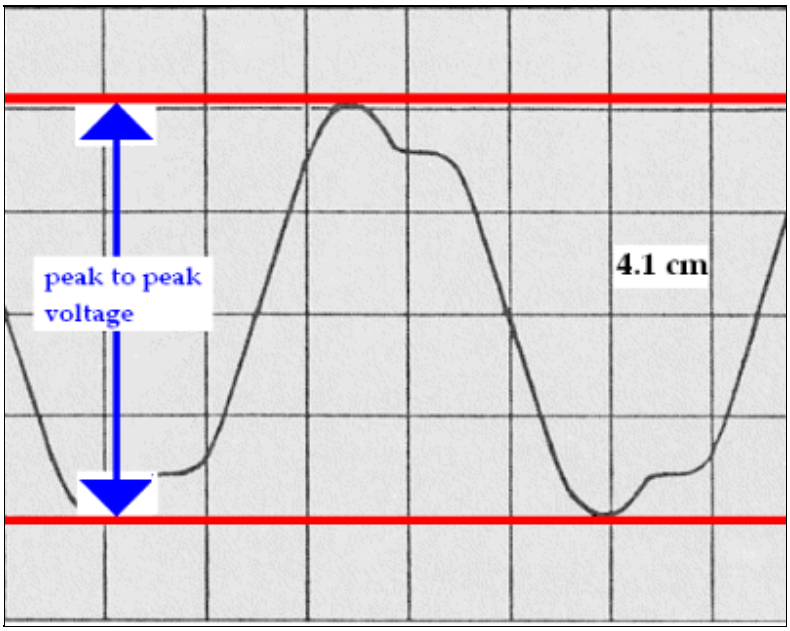
The first installment of a multi-part series exploring basic 50 Ω RF circuit measurement. This web series borrows heavily from the work of Wes, W7ZOI.

RF Workbench 1 shows a way to measure AC signals and quantify power. As circuit builders, knowing the power gain *or loss* in dB of stages like filters or amplifiers; or the absolute power in dBm of our RF signal sources dominates our bench work.

Ignite your bench measurement — Better measurement fidelity inspires confidence, creativity and fun!

Calculate Power Gain from your Oscilloscope Peak-Peak Voltage

An oscilloscope peak-to-peak voltage provides a popular way to determine power gain or absolute power.



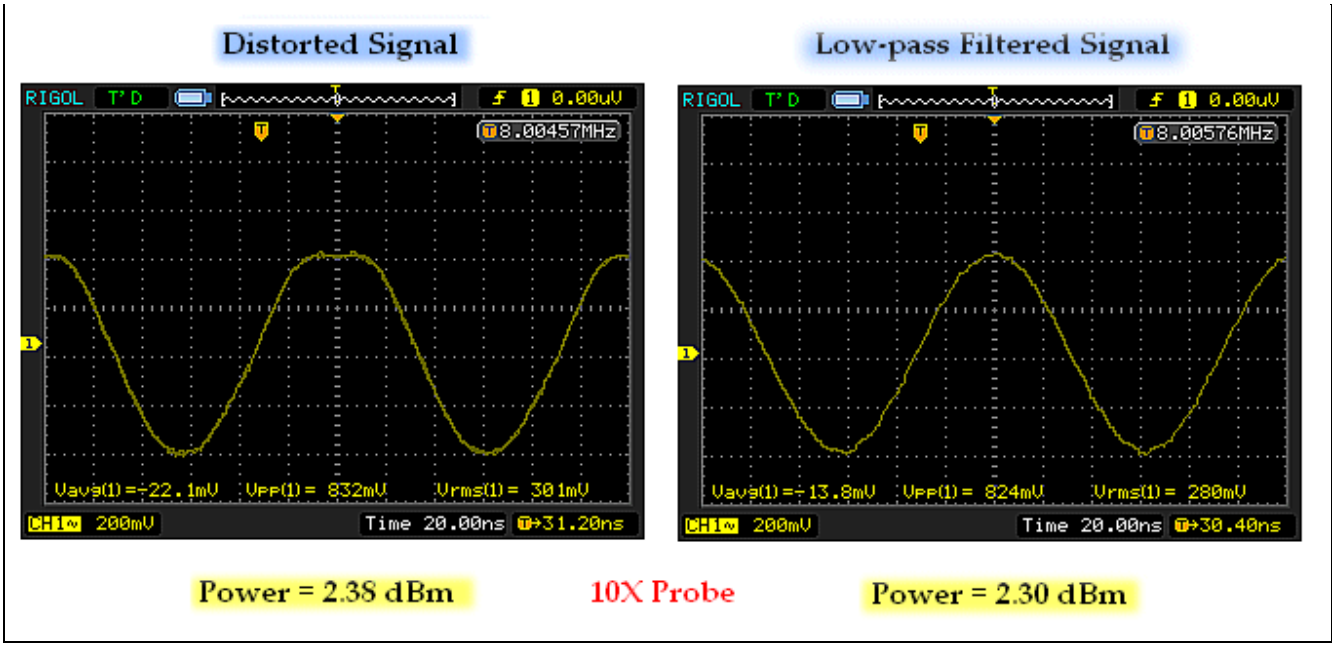
To calculate the power of a sinusoidal waveform, you measure its AC voltage. Assuming your 'scope is calibrated, the first step is to measure the amount of vertical deflection on the screen.

In the figure shown left, the vertical deflection = 4.1 cm. Multiply this measurement by the volts/cm setting of your scope. Lets assume your scope was set to 0.1 volts/cm. Thus the result = 0.41 volts. One final multiplication is required; you must multiply the resultant voltage by the attenuation ratio of the probe. In most cases, a 10X probe is used. Therefore, the measured peak-to-peak voltage is 0.41 x 10 = 4.1 volts. DSOs output numeric voltage readings in addition to that shown — a nice feature.

This signal exhibits a major problem; it's distorted. To calculate the power from peak-to-peak voltage, you require a sine wave. To obtain a sine wave, this distorted signal must be low-pass filtered. Let's examine this topic with some real experiments.

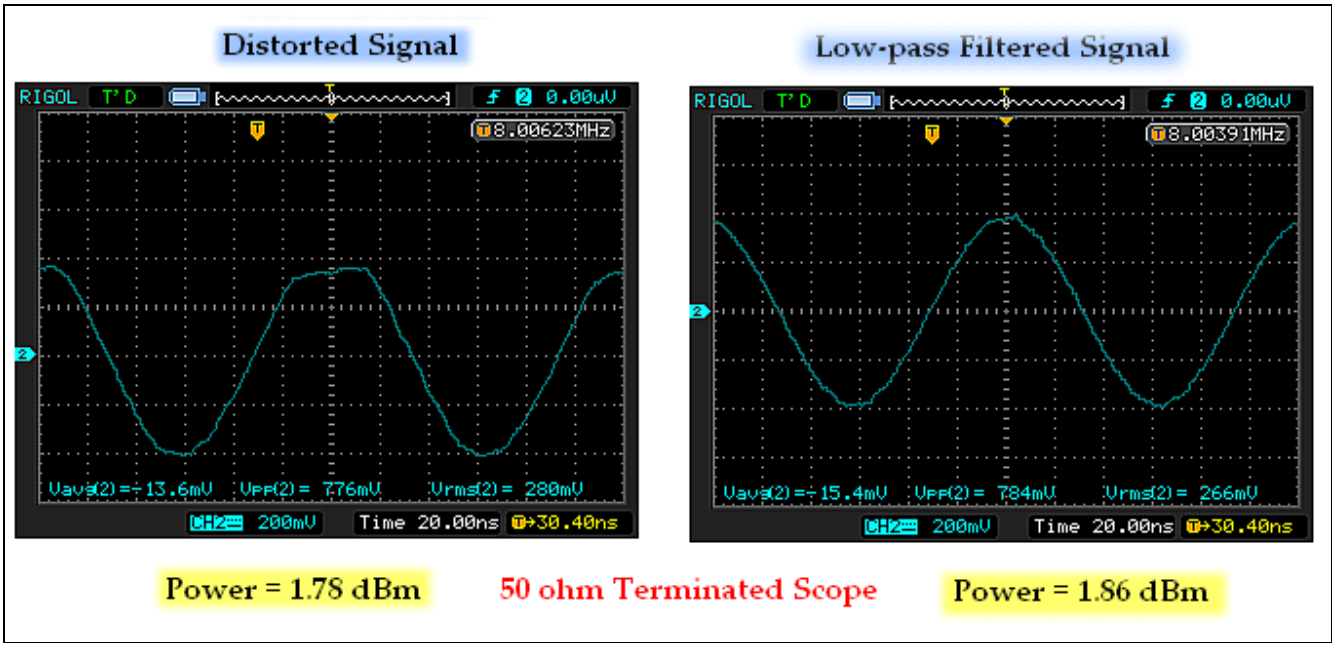
Signal Viewing versus Power Experiments

Measuring 'scope signals takes some skill to get accurate, reproducible results. Consider the following signals taken from an 8 MHz VFO.



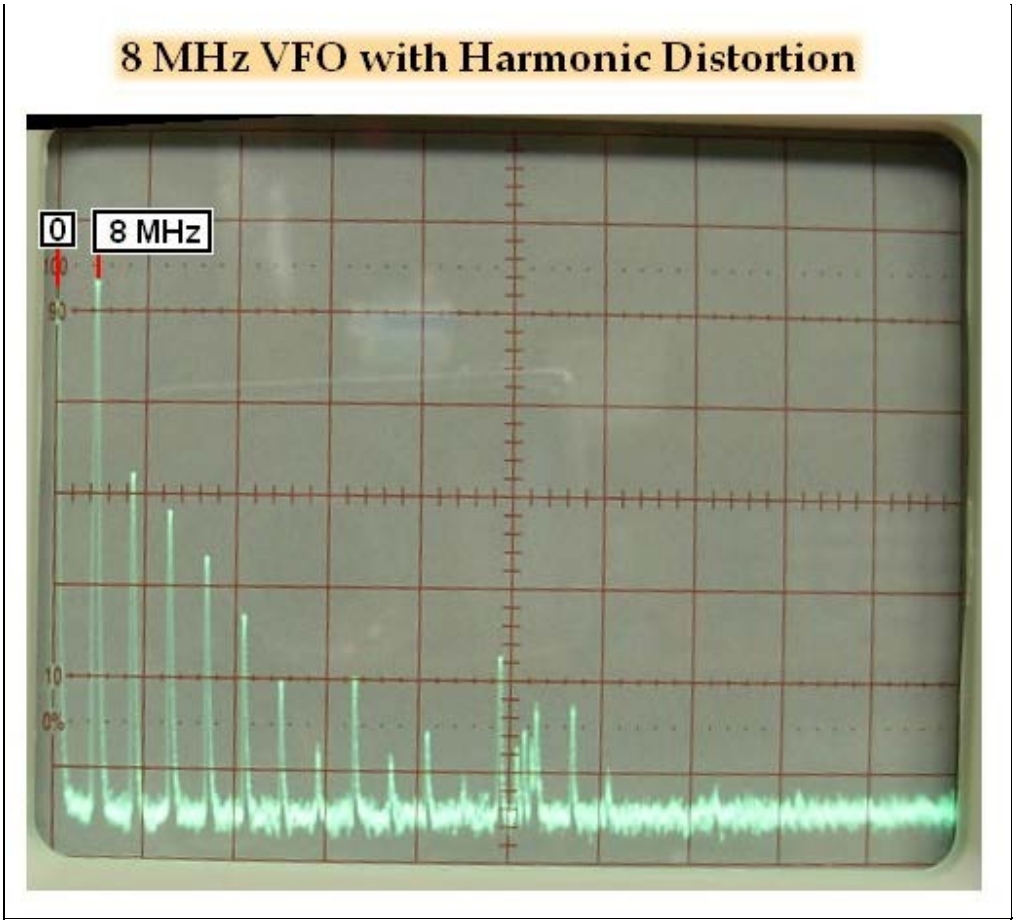
Examine the distorted signal on the left and compare it to same signal after low-pass filtering. I calculated the power in dBm from the peak-peak voltages (Vpp) shown as 832 mV and 824 mV: left to right respectively.

Only the sinusoidal power proves accurate. I'll discuss the formula to assess power soon.

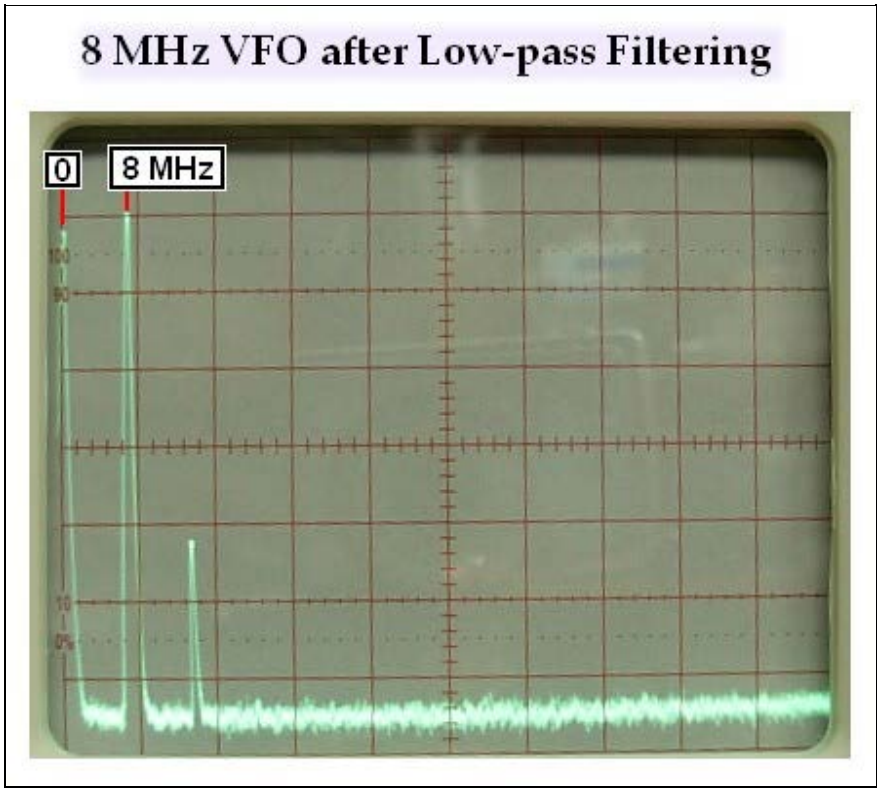


The same 8 MHz VFO examined with a 50 Ω terminated oscilloscope; a superb measurement technique that offers greater sensitivity. In all 4 cases shown, the vertical scale = 0.2 volts/cm.

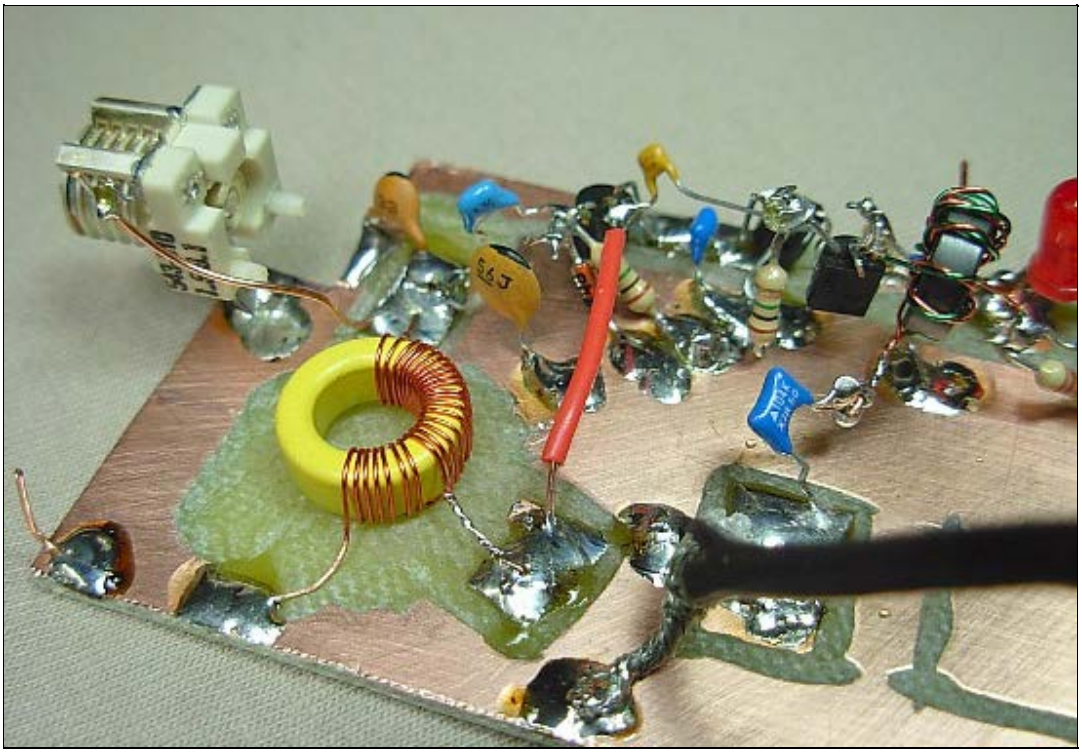
To better compare the distorted and filtered signals, I attenuated the output of the VFO to allow safe examination with a spectrum analyzer. A spectrum analyzer graphs the power (in dBm) of all measured frequencies on its Y axis against a user defined frequency range on the X axis.



Spectral analysis of the distorted 8 MHz signal. The second harmonic (16 MHz) is about 22 dB down from the fundamental. The signal is rich in harmonics that causes error in the calculation of the output power. Each vertical square is 10 dB. Each horizontal square = 20 MHz. The harmonics go 2x fundamental, 3x fundamental, 4x fundamental and so on.



Spectral analysis of the 8 MHz VFO after passing through an N = 5 Chebyshev low-pass filter. The second harmonic now lies about 40 dB down from the 8 MHz carrier (-40 dBc) and the 3rd harmonic is almost down in the noise. Each horizontal square = 10 MHz.

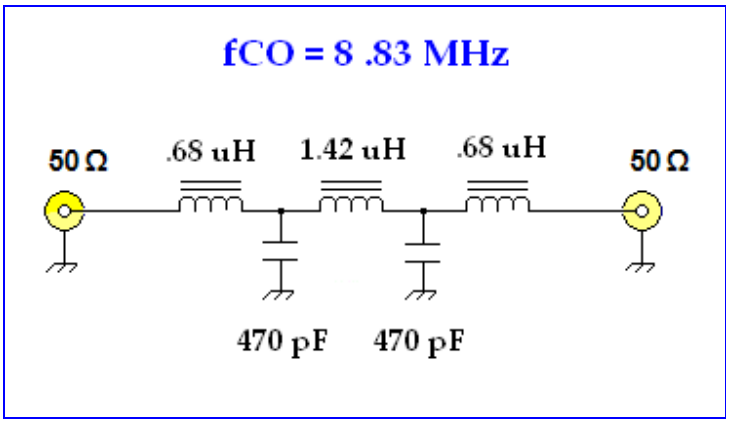


The breadboard of the 8 MHz oscillator oscillator from the above experiments. The output drove a BJT amp biased to give distortion and a 50 Ω output. I adjusted the frequency with the high Q air-variable trimmer capacitor seen to the left.

Low-pass Filter

Some builders wonder why I only employ sine wave signal generators on this web site. To calculate power, they require no low-pass filtering — now you know why. If you're calculating power from a distorted signal, a stiff low pass filter helps ensure measurement fidelity. All of my signal sources feature a 2nd harmonic response of at least -30 dBc, but -50 dBc is typical. To filter receiver front ends, signal generators, or mixer outputs, I keep several 7 element low-pass filter bench modules on hand that cover several 3 dB cut-off frequencies between 3 and 60 MHz.

Although, any old low-pass filters might work fine, Wes, W7ZOI suggests an N = 5 Chebyshev with 0.2 dB of ripple at about 1.2x the signal frequency as a starting point for designing a test-bench low-pass filter. If you don't know how to design low-pass filters choose a pre-designed filter from a filter table. For the experiments above, I selected a filter from an ARRL Handbook. See the schematic below:



Calculating Power (dBm and mW) from Peak-to-Peak Voltage

To calculate the power from peak-peak voltage, the load impedance (Z) must be known. In RF design, the standardized impedance value = 50 Ω. For CATV and video, 75 Ω is common, and in audio and telecommunication design, a 600 Ω impedance dominates. Although we can technically employ any Z, this web site conforms to the 50 Ω RF impedance standard.

The SI unit of power is the watt. In radio, we might see the term dB used, however, dB is a decibel comparison between 2 signals and not an absolute value like the watt. On the bench, dBm serves as the most common and useful term — dBm is the measured power ratio in decibels

referenced to 1 milliwatt.

dBm represents an absolute power — so useful because both large and small signals are quantified with 1 number. Some important bullets follow:

- 0 dBm = 1 mW
- 3 dBm = ~2mW; so doubling the power from 0 dBm equals a 3 dB increase in power
- Increasing the power from 0 dBm to 10 dBm boosts power by 10 dB. The power is now 10X baseline or 10 mW
- 20 dBm = 100 mW
- -27 dBm means that the output has ~500 times less power than 1 milliwatt. -27 dBm = 0.002 mW or 2 microwatts

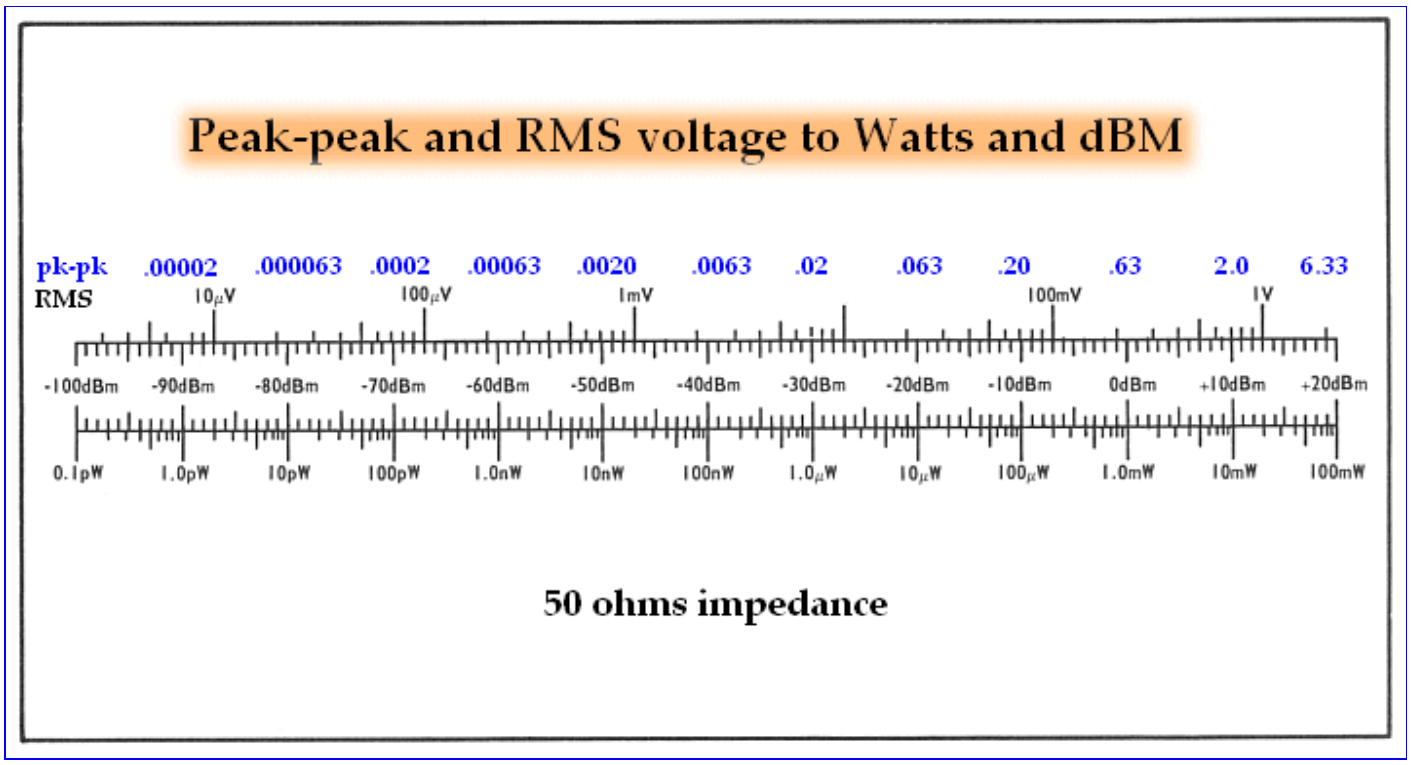
Hopefully over time, you' ll ingrain the concept of logarithmic power gain or loss (in dB) and power referenced to 1 mW (in dBm). This is bread and butter radio design information you must know.

50 Ω Measurement Virtues

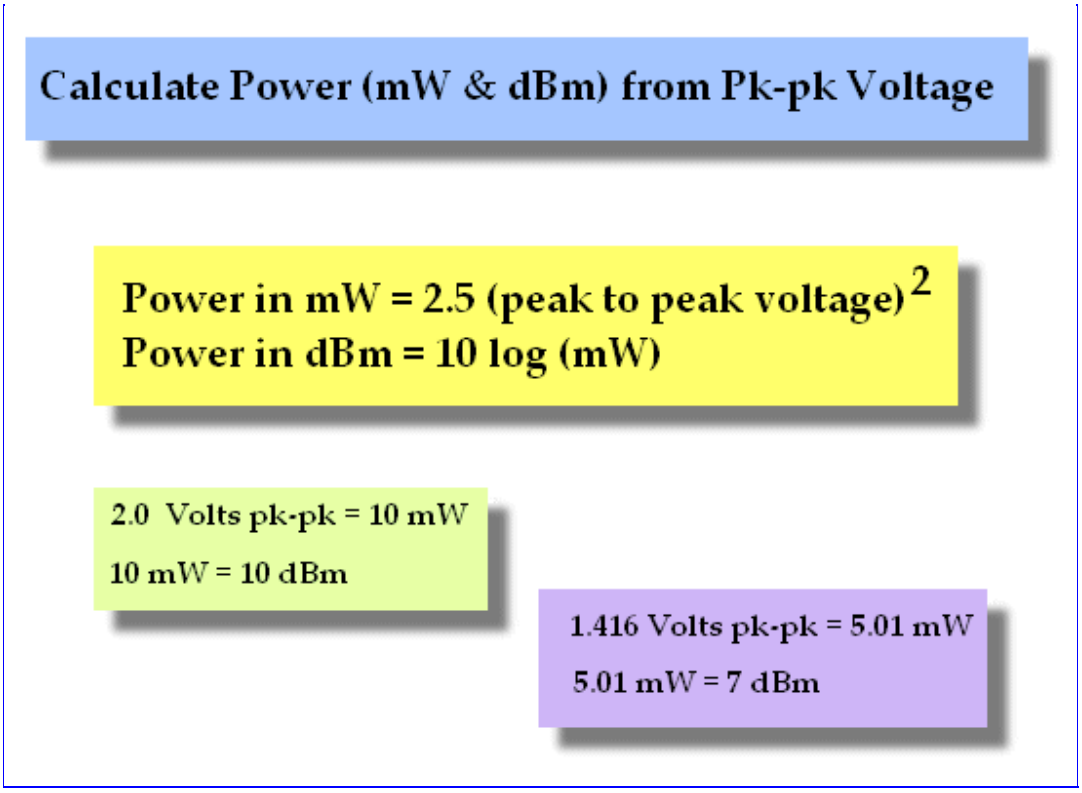
You build a VFO, measure it with your 'scope; calculate the output power into a 49.9 Ω resistor and then record this power in dBm. Let's say it's 6 dBm. VFO output power = 6 dBm.

Next, you place a 6 dB attenuator pad on the VFO output. VFO output power now = 0 dBm.

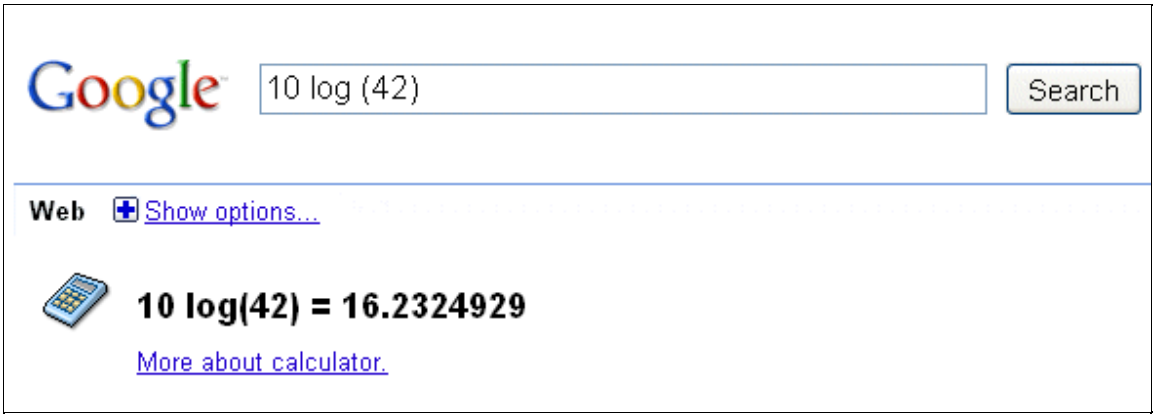
Finally you connect a 10 dB gain RF amplifer to your VFO. Your VFO output now = 10 dBm. What a beautiful system ! ..... it really gets fun when we measure down at - 30 dBm and so forth.



The chart above really helps you visualize the relationships of mW, AC voltages and dBm



In order to get from peak-to-peak voltage to power, math is required. I show the formula above. You may elect to skip the math and calculate dBm or mW from peak-to-peak voltage with software. A number of programs are available; I wrote 1 here as Applet **F**.



If you lack a scientific calculator, Google has math functions. Shown above = a logarithm calculation

In Search of 50 Ohms

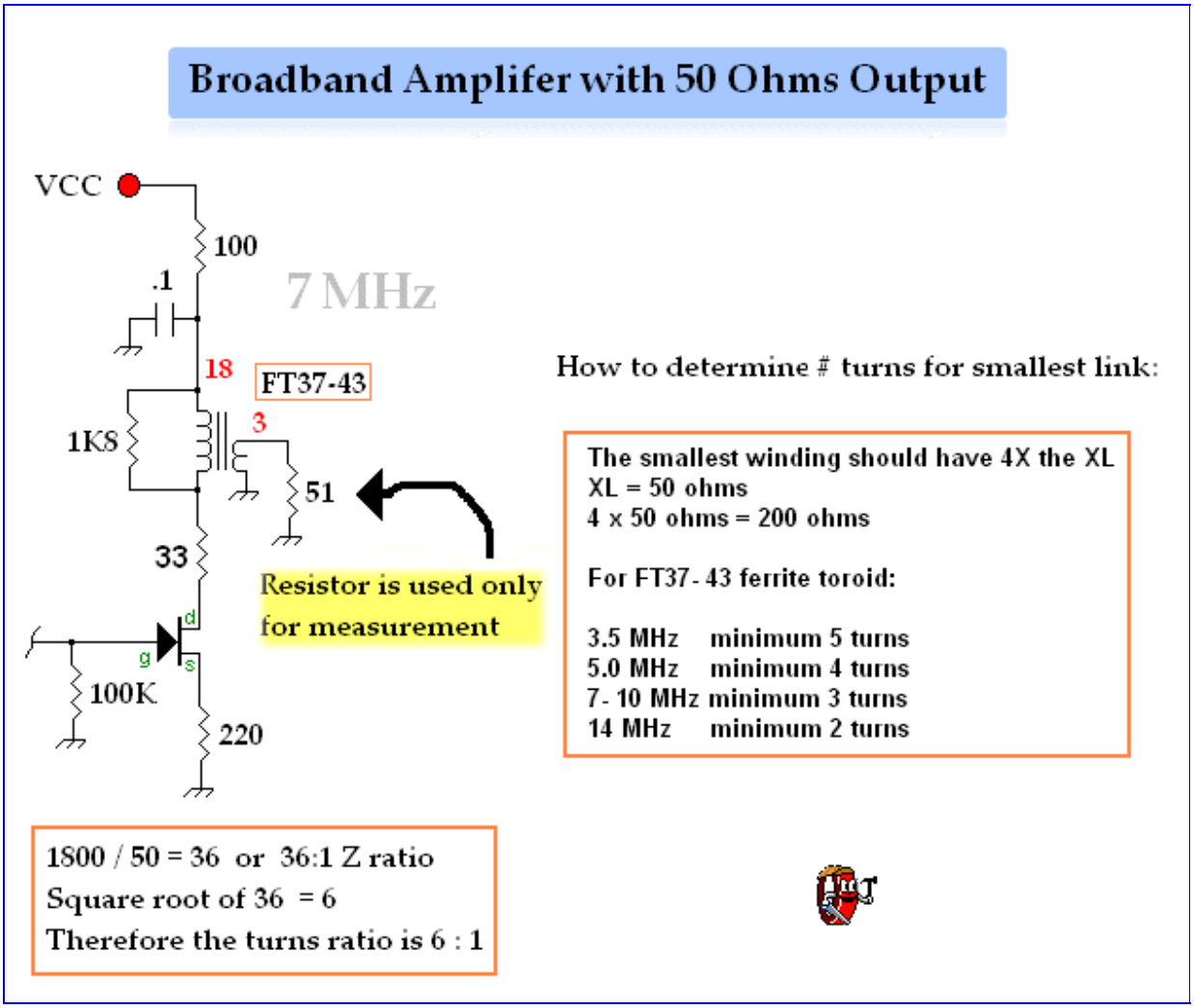
How do I establish a 50 Ω output impedance in my RF amplifier? I get this question often.

From my experience, in **simple** amplifiers lacking negative feedback, a 50 Ω output impedance must be created by inserting a resistor somewhere in the circuit that forces a 50 Ω output impedance.

For example, we might use a 50 Ω collector resistor in a BJT amplifier (or a 50 Ω drain resistor in a JFET amplifier), or place a fixed resistor in parallel with the collector/drain transformer and use a secondary winding to establish the 50 Ω output impedance. Sometimes we'll place a series resistor (say from 22 to 51 ohms) on the output of an emitter or source follower to bring the low output impedance up to 50 Ω. These form basic explanations and usage examples may be found on many schematics on the QRP / SWL website.

We experimenters also employ negative **feedback** with or without output transformers to establish a 50 Ω input and/or output Z and I show many examples on this site.

The following diagram explores 1 method to get a 50 Ω output impedance in a simple amplifier. It doesn't matter if the transistor is a JFET or a BJT, the principle is the same. This diagram and tutorial are simplistic and meant to help novice builders learn to design their own amplifier stages. You may connect any resistor value across the output of a transformer to calculate power, however, this web site only considers 50 ohms.



The above diagram describes a broadband (untuned) amplifier. I employed a FT37-43 ferrite toroid: a common part. Other ferrite toroids may be substituted, however the table depicting the minimum number of turns won't apply.

Consider the BJT amp shown. The transformer primary winding is shunted with a parallel 1800 Ω resistor. The 1K8 resistor "forces" a 1K8 ohm collector output resistance in the primary winding.

To transform the 1800 Ω primary impedance to 50 ohms; use a 3 turn secondary link. Calculate the primary to secondary turns ratio as follows:

1800 ohms divided by 50 ohms = 36. The impedance ratio = 36:1.
The turns ratio is the square root of the impedance ratio; thus the turns ratio is 6:1. The primary winding must have 6X the number of turns of the secondary winding. In the 3rd RF Workbench web page, you'll see that the above explanation pertains to the "ideal transformer", however, the concept is useful — especially to the target audience of this website.

New builders might ask — why not wind 6 turns for the primary winding and 1 turn for the secondary winding? We avoid this because the smaller or secondary winding should have have a minimum inductive reactance (XL) of 4X the impedance it is connected to. Thus for a 50 ohm circuit, the minimum XL = 200 Ω at the design frequency.

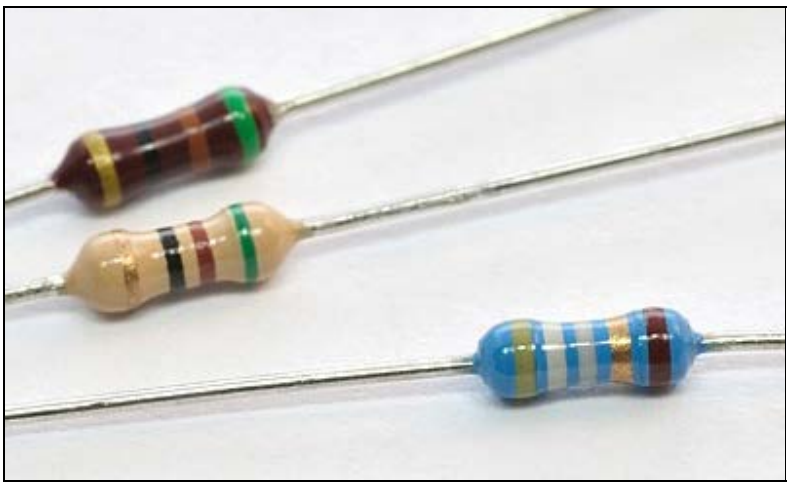
This design rule serves only as a rough guide. We employ the minimum 4X rule because employing an XL less than 4X may create unwanted signal losses and affect the smaller winding's impedance. The table to the right of the amplifier shows the minimum numbers of secondary turns for a few common frequencies with the FT37-43 ferrite toroid.

Thus for our 7 MHz amplifier, we need at least 3 secondary turns and multiply this number by the turns ratio to give a 18:3 turns ratio. You might also choose 24:4.

For ferrites other than the FT37-43, calculate the minimum number of turns with the XL= 2 pi * F*L formula and detemine L from the turns versus AL toroid data, or measure L with an inductance meter.

For AC measurement a 50 Ω purely resistive load should be temporarily connected between the output link and ground. This might be a 51 ohm resistor, a 49.9 ohm 1% metal film resistor, 2 parallel 100 ohm resistors, or some other "50 Ω" load. We measure peak-to-peak voltage across

the load and then calculate the power in dBm or mW. After measurement, the temporary 50 Ω load is removed and the circuit connected to the succeeding stage.

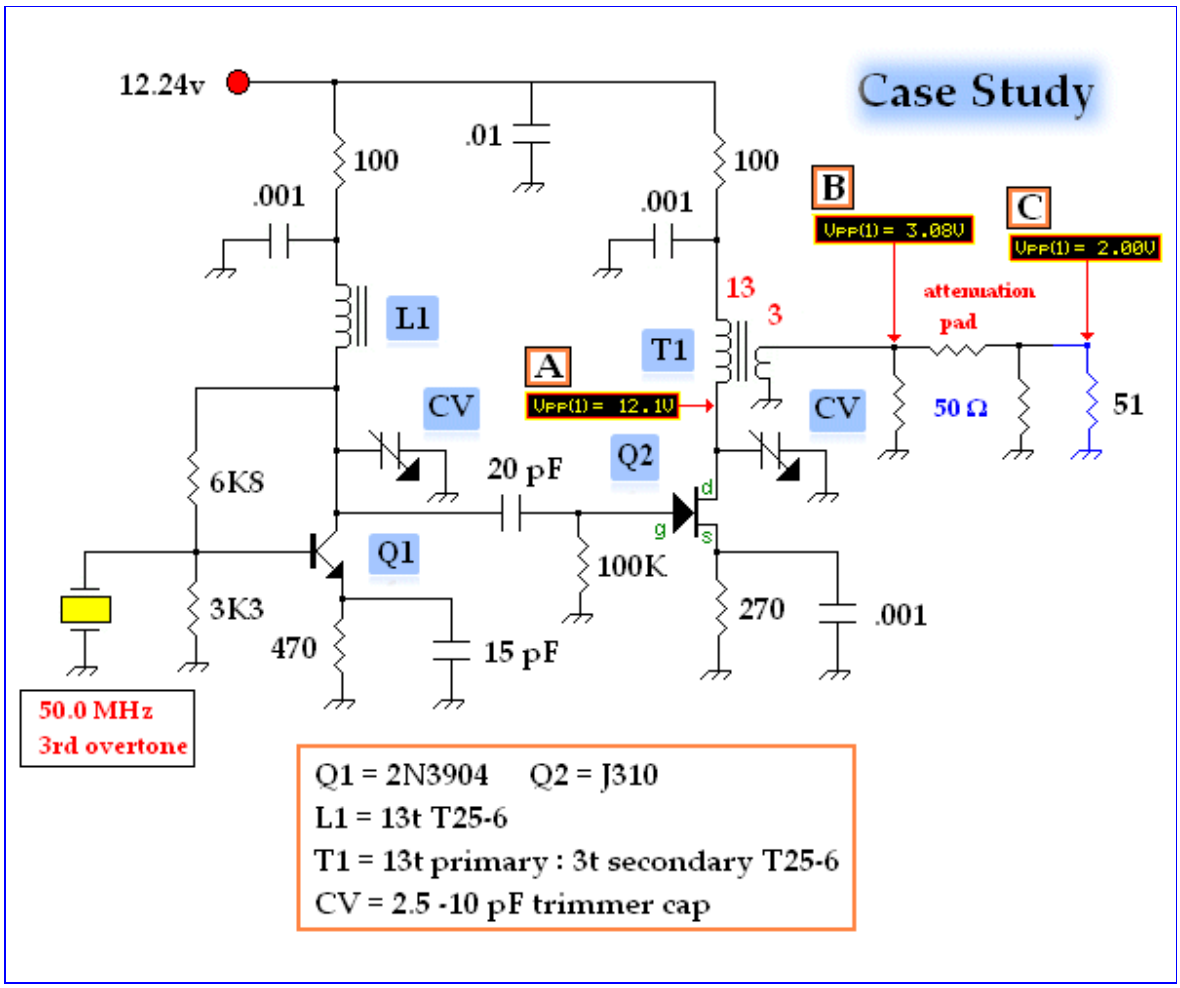


51 and 49.9 Ω resistors

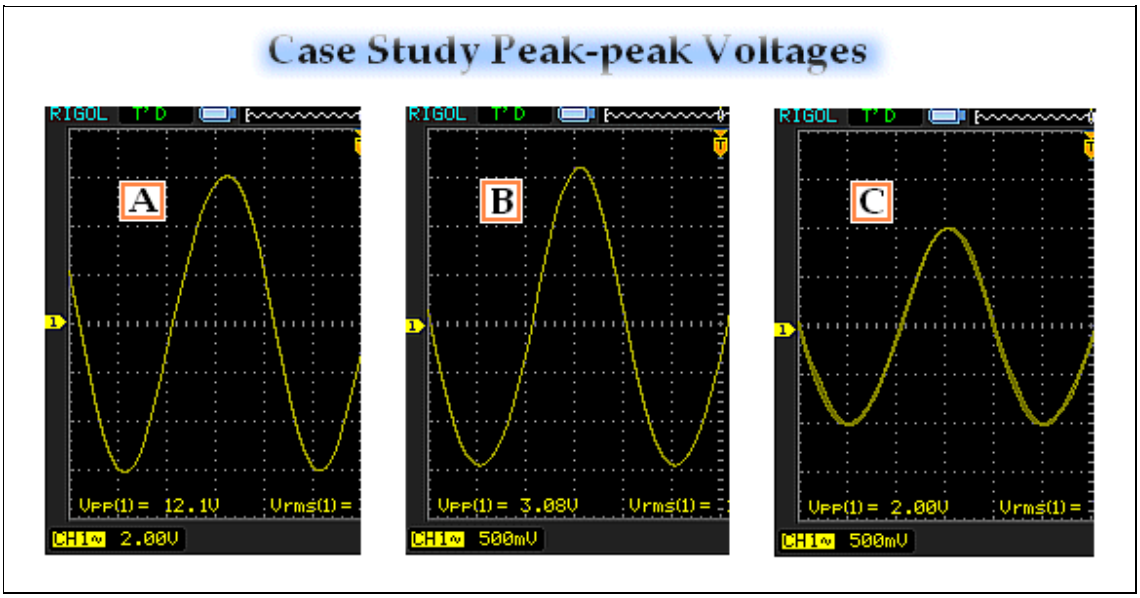
We measure the peak-peak voltage across this 50 ohm resistor with a 10X 'scope probe; or alternately may connect the device output to a 50 Ω terminated scope to measure peak-peak voltage to calculate power.

I normally measure in a 50 Ω measurement environment and temporarily solder a BNC connector onto my breadboard and connect this port to a 50 Ω terminated scope with coax. After testing and voltage measurement, I remove the RF connector and then build and test the next stage.

Case Study



Pretend that you breadboarded the above circuit entitled "Case Study". This is a 50 MHz crystal oscillator and buffer. The crystal fundamental frequency = 16.7 MHz, but the L1 tank is tuned to its 3rd overtone; 50.0 MHz. You measure and record the peak-to-peak voltages at the points labeled A, B and C.



The peak-to-peak voltages are shown as Vpp. The vertical scale (volts/cm) is shown on the bottom of each figure.

Examine Point A. The AC voltage = 12.1 volts peak-peak. Compare this to the peak-peak voltage at Point B. Note the difference. Some builders emailed me after they measured similar differences on the primary and secondary transformer windings of their circuits with a 10X probe. These builders felt something must be wrong. All normal; you can expect the peak to peak voltage to roughly decrease (or increase) by the transformer turns ratio.

The 12.1 volts peak-peak decreased by a factor of 4.3 (13 / 3 turns ratio) which is 2.8 volts peak-peak. In our case, the measured secondary peak-peak voltage was 3.08 volts — in the ballpark. Please remember this serves as a **coarse** guide only. It helps you to know what to reasonably expect during signal viewing.

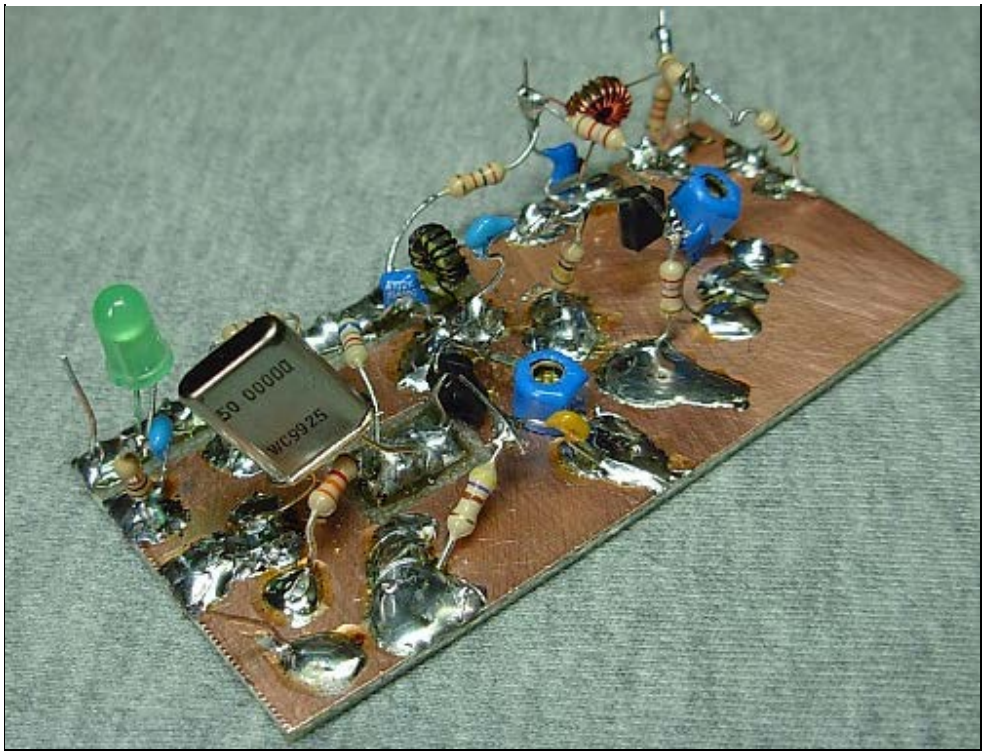
Peak-to-peak voltage changing in accordance with the transformer turns ratio represents a simplistic explanation describing the "ideal transformer". To understand real world transformer function, you must contemplate factors such as Ohm's law for AC, conservation of energy (this is what causes the voltage to drop while preserving power) and basic transformer behavior. These principles are explained in publications such as The ARRL Handbook for Radio Communications, or the RSGB Radio Communication Handbook. An old high school physics text book might prove a better reference.

Here are the case questions:

1. Calculate the power in dBm at point B
2. Calculate the power in dBm at point C
3. What is the attenuation in dB of the 50 ohms attenuation pad?
4. What is the output power in mW of this stage?

Click on this link for the answers and to see the actual resistor values of the attenuator pad.

Finally, placing a 10X probe at Point **A** will de-tune the L-C tank circuit somewhat and thus alter the AC voltage. In real-world building; to tune Q2, tweak the variable capacitor (CV) with your 10X probe connected to Point C.



The breadboard of the 50 MHz oscillator prototype.

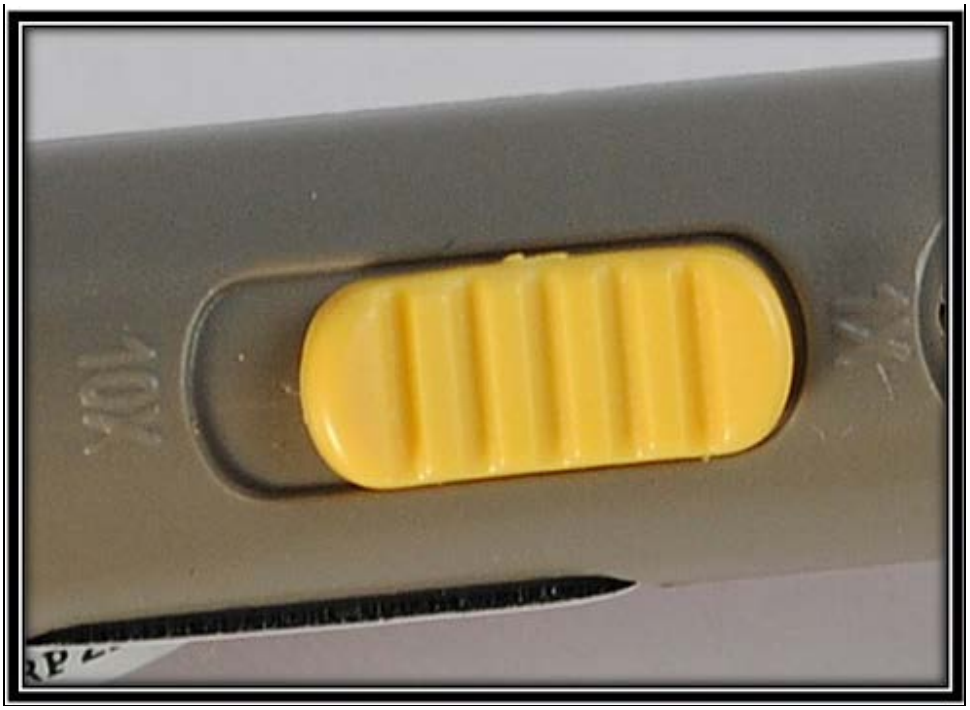
Oscilloscope Probing

10X Oscilloscope Probe

Please refer to EMRFD Chapter 7 for great information about measuring power in RF circuits. The 10X oscilloscope probe is one of the most important measurement tools to have on your bench. There are countless web articles concerning the 10X probe, so I don't have much to add.

Take care of your 10X probe: don't solder components you've clipped your probe to; avoid setting heavy objects on the cable; store it carefully and inspect it frequently.

When do you use a 10X probe ? Measure with a 10X probe for in-situ ("in place", or "in circuit") voltage measurement and in situations where you can afford a 10X reduction in sensitivity. In low level measurements such as millivolt level measurements, the reduced sensitivity of a 10X probe may reduce or disallow accurate measurement. Additionally, the 20 pF or so capacitance of a 10X probe can detune resonant circuits; especially at VHF on up.



Close up of the Rigol oscilloscope probe 10X and 1X switch.

50 ohm Terminated Oscilloscope

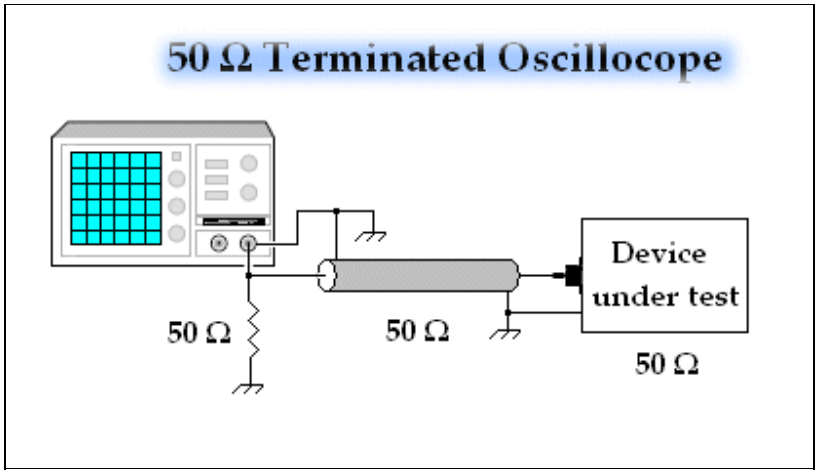
At RF, we generally work with (or try to work with) circuits with 50 Ω impedances. If possible consider performing your measurements in a purely 50 ohm environment.

That is — instead of using a 10X probe, shunt the oscilloscope input port to ground through a 50 Ω resistor and connect your test circuit to the 'scope with 50 ohm impedance coaxial cable. On my 'scopes, I have Channel 1 set up for the 10X probe work and Channel 2 set up for a 50 ohm environment.

I asked Wes about the benefits of performing measurement in a 50 Ω environment. I learned the main advantage of a 50 ohm approach is a well defined port impedance. The second virtue; a 10X greater voltage sensitivity — the increased sensitivity for low level measurement amazes me. In some cases, small signals that I couldn't accurately measure with a 10X probe, gave an excellent scope tracing with more consistent voltage readings in a 50 Ω environment.

You also may enjoy improved signal viewing. For example, in a few cases I have observed harmonic distortion with a 50 ohm terminated scope unseen with a 10X probe I confirmed this distortion with a spectrum analyzer.

If you have never performed measurement in a 50 Ω environment, consider trying it out — you'll enjoy it. You may buy commercial 50 Ω feed-through devices that connect to your oscilloscope input, or homebrew your own, but try to keep the 50 ohm termination as close to the oscilloscope input as possible.



Establishing a 50 Ω impedance measuring environment. The oscilloscope input is terminated with a 50 Ω resistance and connected to a device with a 50 Ω output impedance via 50 Ω coaxial cable.

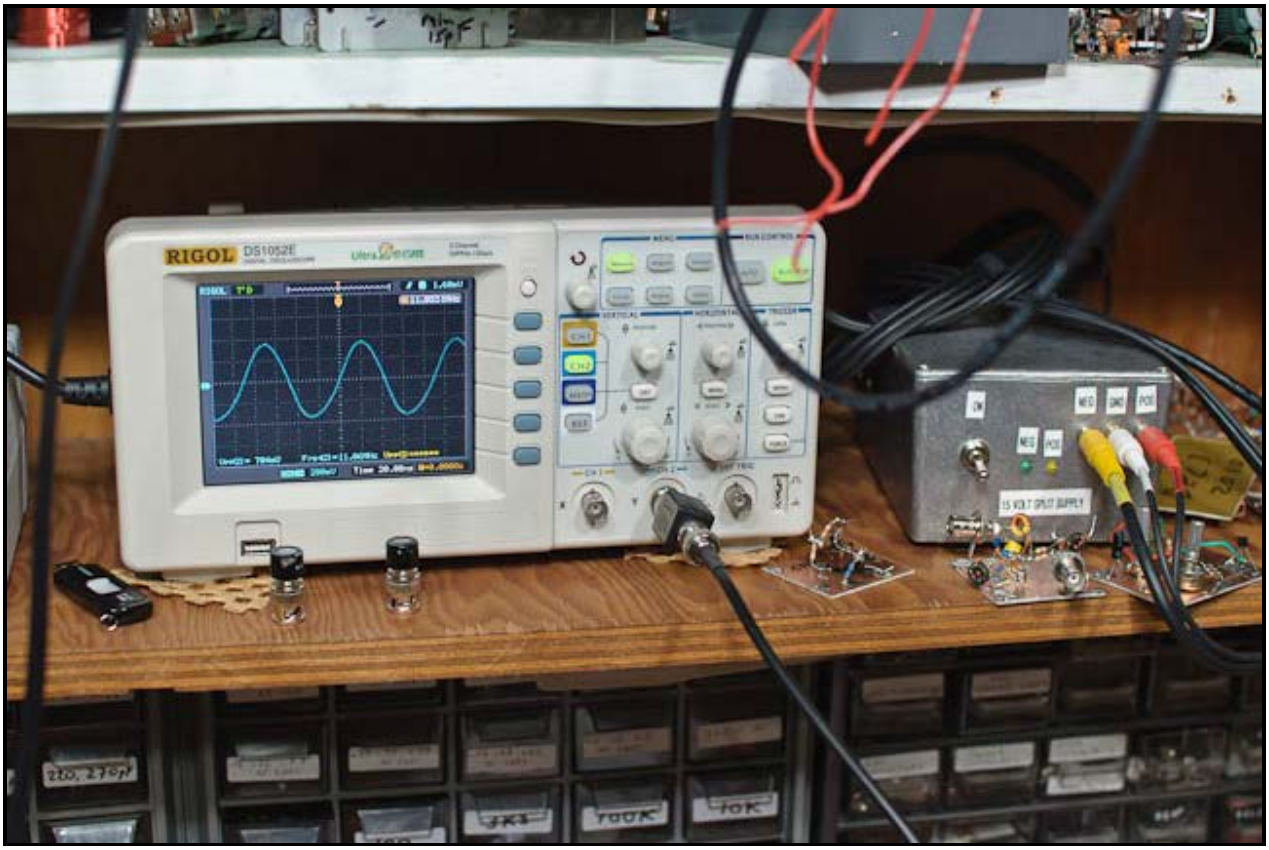
Try not to routinely connect a feed-through attenuator pad to your feedthrough 50 Ω 'scope terminator — error may arise.

My very first homebrew 50 Ω scope terminator module with 2 female connectors. I connected this module to my oscilloscope input via a commerical 9 cm long 50 Ω coaxial cable with a male connector on each end.

Two parallel 100 ohm resistors formed the 50 Ω load. Ideally, the 50 Ω shunt resistor should be right at the 'scope's female BNC jack — so this homebrew module shown fell short as a stalwart 50 Ω terminator. Inspired to move to a 50 Ω environment and lacking a male BNC connector, it did the job until my commercial version came by mail. You might find oscilloscope feedthrough terminators for sale at Ham festivals.

An ideal homebrew solution — place a male and female BNC connector in a small metal box very close together to allow a very short interconnecting wire. The box would hang off of the oscilloscope. Better still are commercial, shielded 50 Ω feed through terminators which thread right onto the oscilloscope's female BNC input jack.

Above — a commercial 50 Ω feed-through BNC terminator on my oscilloscope input.



Above — Measuring in a 50 Ω environment. Bliss! Я люблю это.

RF Current Sampler

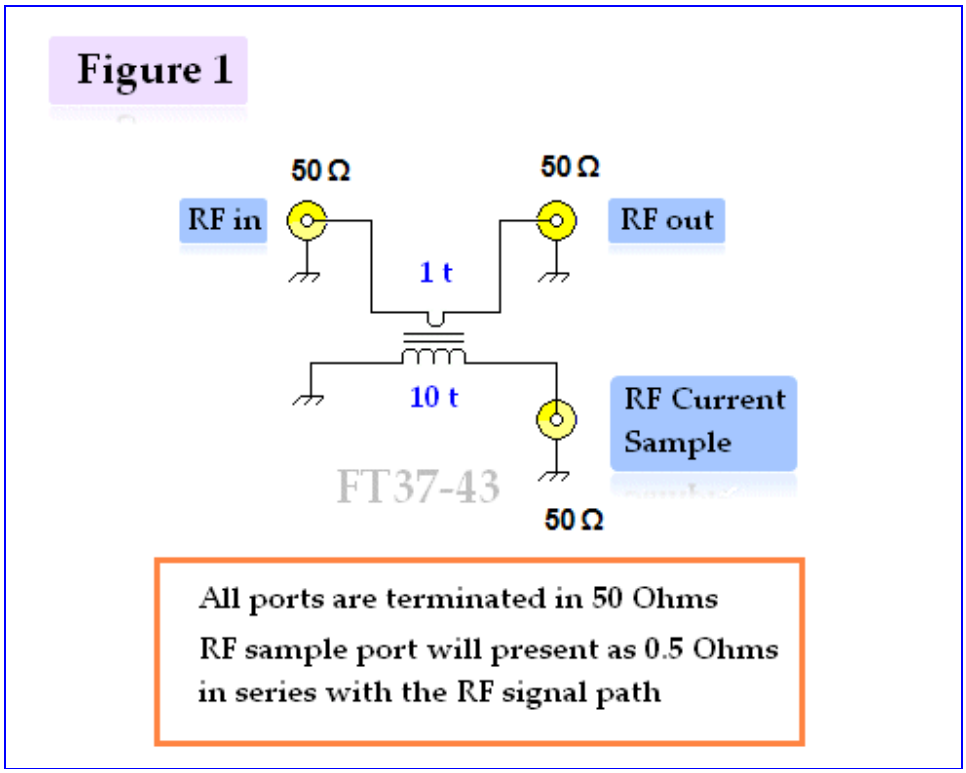
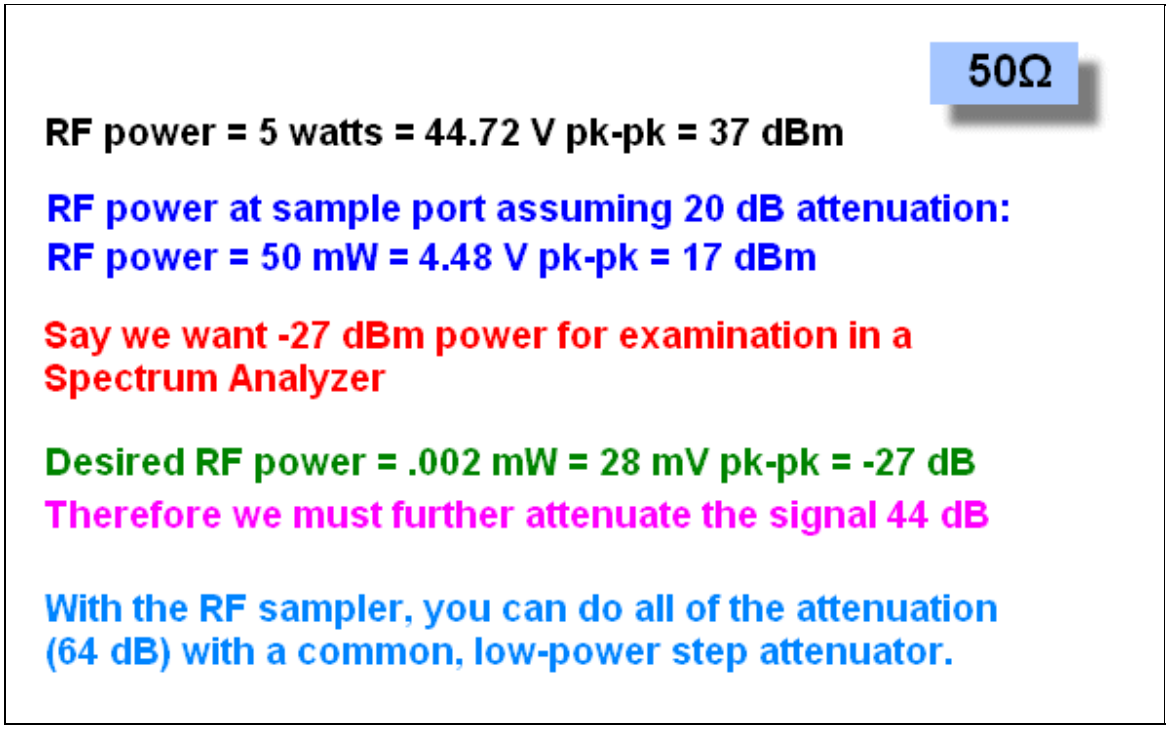


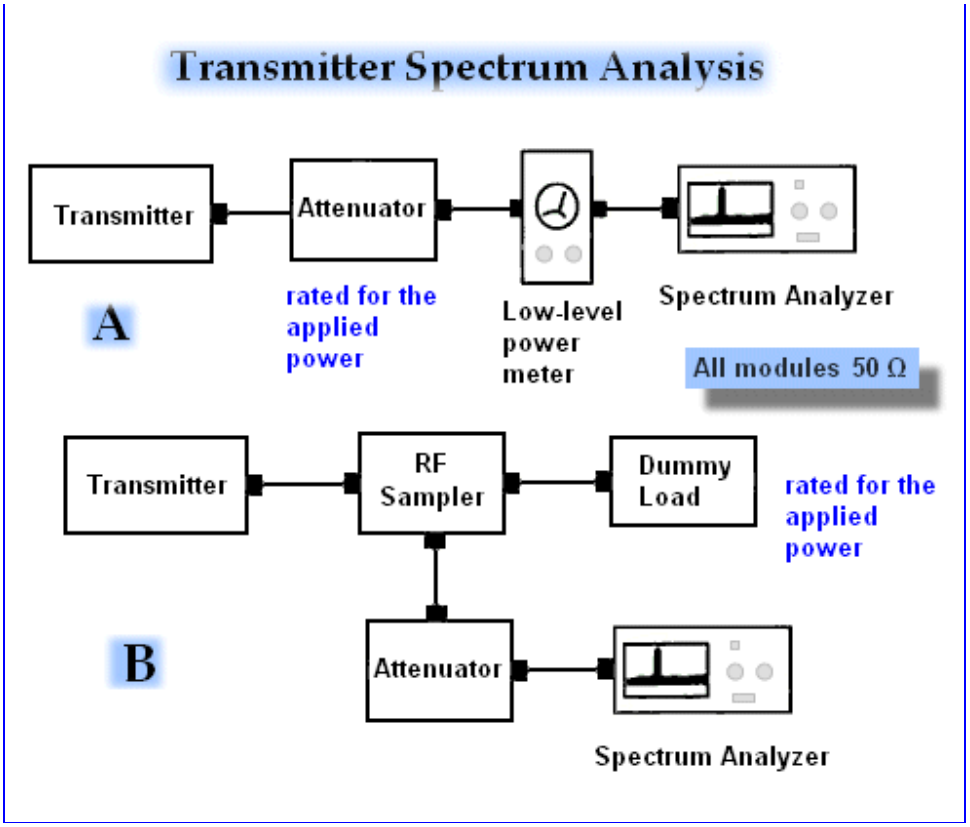
Figure 1 shows a basic circuit to sample RF current from a power stage such as a QRP transmitter. Many experimenters lack 50 Ω step attenuators rated to handle transmitter-level power. One basic solution is to sample the RF current of the power amplifier using a wideband step-

down transformer. Terminate the RF current sample port with a 50 Ω impedance device. This may include a spectrum analyzer, power meter, receiver with an attenuator, or a 50 Ω terminated oscilloscope.
A usage example = examining a transmitter's spectral purity with a spectrum analyzer. The output power at the sample port will drop by 20-22 dB. A 50 Ω impedance step attenuator can be used to further reduce this power level to whatever you want. For this chore, a typical experimenter's 1-2 watt step attenuator works, since it never "sees" the higher wattage transmitter power.

For example, a 5 watt amplifier 20 dB down is 0.05 watts or 50 mW at the RF current sampler port. 50 mW = 17 dBm. To examine this signal with a spectrum analyzer you may wish to decrease the power down to -27 dBm. The following chart shows the basic process.



A Hammond chassis shields the RF current sampler used on my bench.



The above graphic illustrates 2 methods to examine the output of a transmitter in a spectrum analyzer. Method B is described above. The dummy load must handle the transmitter output power, however a 5 or 10 watt dummy load is easy to make. Method A requires a step-attenuator which can handle the transmitter output power. The low-level power meter promotes the need to quantify the output power before you connect anything to the output of the attenuator. This is also true for Method B.

Always measure the output power at the RF sample port with your oscilloscope or low-level power meter before hooking up any expensive low-level measurement device such as a spectrum analyzer! осторожно!

Miscellaneous Photos

Above — a 50 Ω BNC terminator. These are essential QRP work bench items and may be found on eBay for cheap.

Click for a photograph of 4 of my BNC RF port terminators: 27, 50, 75 and 100 ohms from left to right — useful to calibrate and test RF circuitry.

RF Workbench Page 2

Welcome to part 2 of a web series exploring basic RF measurement and bench practices. This installment builds on the information from the RF Workbench Page 1

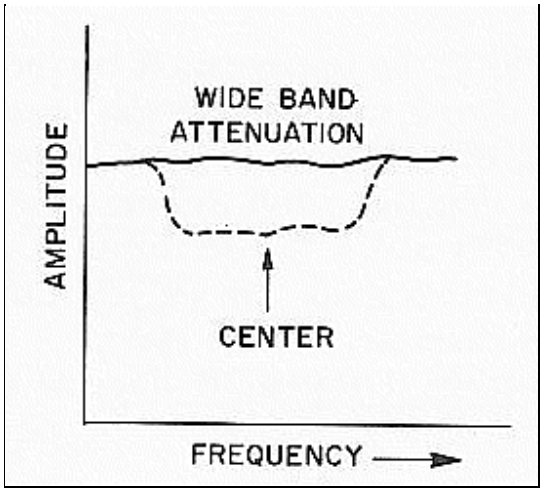
I share introductory and practical content on attenuation, the return loss bridge, insertion loss or gain and spectrum analyzers. Consult EMRFD and use your favorite web search engine for more information.

The Attenuator Network

Like onions in your kitchen, the importance of attenuators can't be overstated. On the bench, attenuator pads go in nearly every test circuit to deliver correct power and/or help match port impedances.

Think of attenuators as passive networks that intentionally insert power loss between 2 components independent of frequency. For example: you might follow a 50 Ω signal source with an attenuator to decrease its power, increase its return loss, and/or buffer it from downstream impedance changes.

- Most attenuator networks have fixed input and output impedances.
- The input and output impedances may be the same, or different.
- Attenuation may be fixed or variable. Most often, we use simple, fixed resistive pads that function as voltage dividers.
- Express attenuation in dB.
- Attenuators increase return loss and reduce VSWR.
- Attenuators may function as buffers to isolate stages.
- All attenuators on this web site feature 50 Ω input and output ports.



The desired response of an attenuator network. Practically speaking, device construction techniques including shielding will limit how high in frequency your attenuator will properly operate.

A pair of commercial 50 Ω input/output impedance step attenuators from the past. Occasionally, you'll find them for sale at Ham festivals or estate sales. Most work well for HF and perhaps even VHF work depending on their design and condition. Visually examine and test the attenuator before use.

How to Design and Build Attenuator Networks

To design attenuators with a 50 Ω input and output impedance, I recommend viewing a table. Click for a table.

After choosing the degree of precision; solder up 2 or 3 resistors and you're done. Nearest value 5% tolerance resistors offer reasonable precision for our popcorn circuits, but combining 1% and 5% tolerance resistors works too.

Choose low inductance resistor such as carbon film types and strive for short lead lengths. Well consider resistor power dissipation — for example, an 8 dB attenuator pad will dissipate 84% of the RF passed through it. I have seen attenuator pads that were exposed to high power and some or all of the resistors were burnt and turned to charcoal. Clearly the operator did not regard the power rating of the attenuator resistors. Refer to EMRFD Section 7.4 for practical information concerning attenuator design and power dissipation.

Three of the attenuator pads from my bench attenuator drawer. When using Ugly Construction or its variants, you can solder in, change, or remove attenuator pads at a whim. A small stock of these pads speeds up your work flow.

A 10 dB attenuator pad from my collection. This box uses two 100 Ω (5%) resistors and a 68 ohm (1%) resistor for the 96.2 and 71.2 ohm

resistances called for. I used the 1% tolerance part because all my 68 Ω resistors are metal film 1% tolerance 1/4 watt types. Perhaps, I'll pursue a closer match to the calculated resistor values 1 day, but this module works okay. You may also stick 2 seperate attenuators in the little Hammond box shown.

Two commercial BNC feed-through (in-line) attenuators. I use these *every* day and prefer them over homebrew R networks since they don't require a coaxial cable. I own many: two 6 dB, two 10 dB, one 3 dB and a 20 dB: all were purchased on eBay.

Two commercial SMA in-line attenuators for my VHF and UHF experiments: 6 and 15 dB. Click for a marker table with 4 data points derived from sweeping these 2 filters.

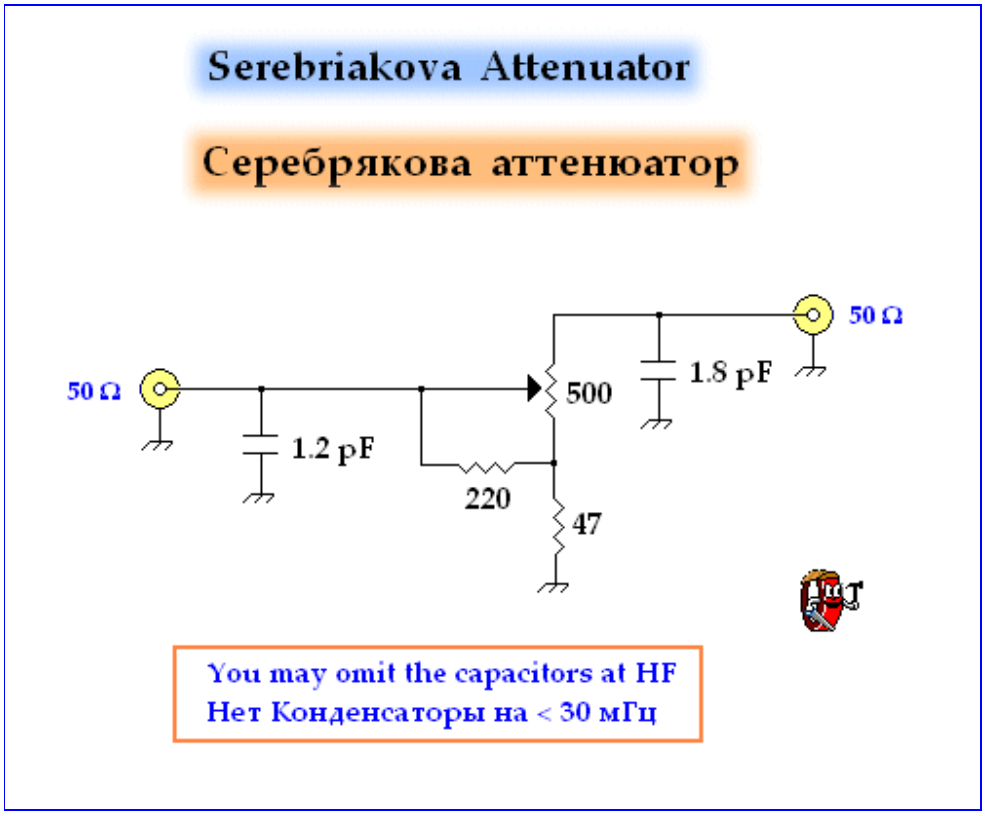
Step Attenuators

A step attenuator belongs on every serious RF workbench. They allow in-situ attenuation adjustment with a degree of precision as low as 1 dB. Step attenuators are nothing more than switched calibrated resistances and the switches can be SPDT, relays, rotary or digitally-controlled types. The quality and price of commercial attenuators varies widely. Experimenter concerns include the minimum attenuator insertion loss, power rating, return loss, noise from switch contacts and noise from the resistors themselves.

A homebrew step attenuator makes a great weekend project and almost every radio handbook contains 1. Web linked projects plus commercial kits may be found online — use your favorite search engine to find them.

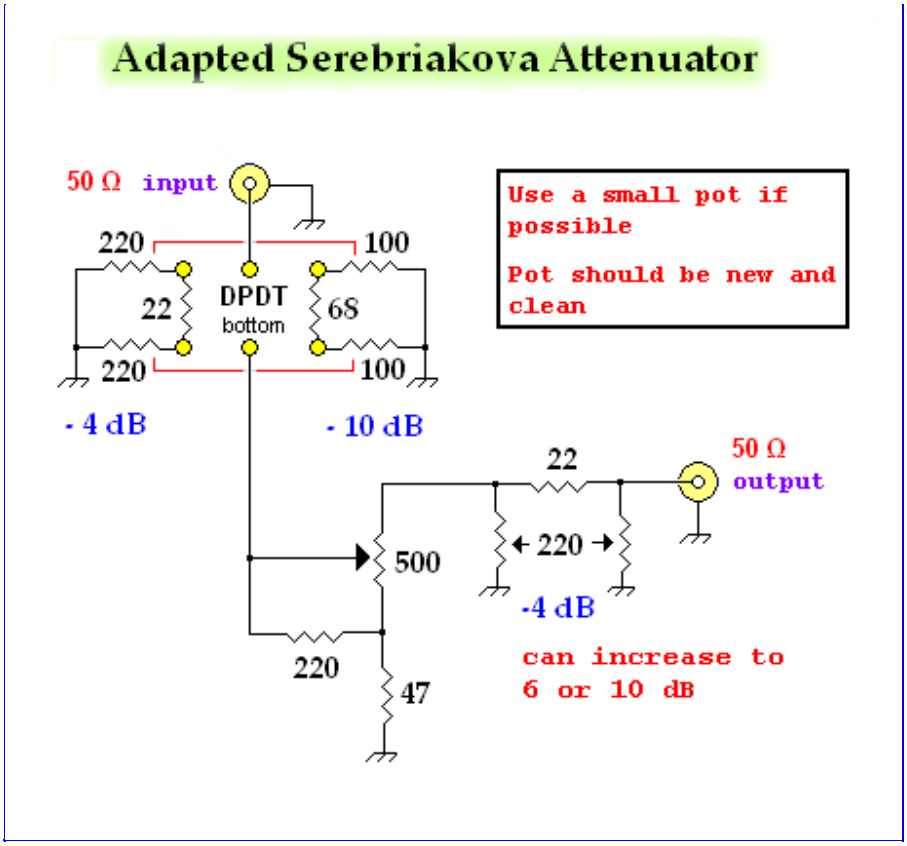
Some homebuilders prefer 1% metal film resistors to keep resistor noise down. Stick your step attenuators in a metal, RF-proof box and insert quality interfaces such as BNC, N or SMA connectors. Your needs, budget and parts collection determine the outcome when you home build one.

Serebriakova Attenuator - Серебрякова аттенюатор (50 Ом)



The Serebriakova; a simple, variable attenuator well suited for QRP homebuilding. Filled with gratitude to its Russian designer's family, I share this contribution with my readers. This attenuator network makes signals smaller or larger in a 50 ohm environment via a potentiometer. My analysis indicates acceptable performance considering its simplicity. The input match is close to 50 Ω across the range of the potentiometer. The output match across the potentiometer range is mediocre. Click here for a DC match analysis from Wes, W7ZOI. As shown, you wouldn't place this device on your main bench signal generator output as the output impedance diverges widely during amplitude adjustment.

Add fixed attenuator pads on the input and/or output to improve matching into 50 ohms. This circuit could serve in multiple applications including an RF gain control on a receiver front end, for bench measurement (when adapted) and for a low-level transmitter gain control. The Serebriakova attenuator may function up to 500 MHz in a carefully constructed, shielded box. The input and output capacitors may be omitted below 30 MHz. The attenuation varies a minimum of 20 dB when turning the potentiometer from CCW to CW. Click for a build by he yl.



A variant of the Serebriakova attenuator is shown above. Input and output matching are enhanced by fixed attenuator pads. The input match into 50 ohms is fine. After testing, I learned that the fixed 4 dB output attenuator pad is likely too low to ensure a wide range output match into 50 ohms. A 6 or 10 dB output pad is preferable, however, if this is your only variable attenuator, the device would then only be usable for very low-level work. You can decide what value of input or output pads to use.

A new, clean and small size 500 ohm pot works best. Store your potentiometer collection in sealed plastic bags to keep out workshop and house dust.

Serebriakova Attenuator RL and VSWR @ 14 MHz			
Setting	Input -4 dB	Input -10 dB	Output
CCW	27.9 dB; 1: 1.08	30.6 dB; 1: 1.06	9.6 dB; 1: 2.00
Midrange	17.08 dB; 1 :1.30	25.2 dB; 1: 1.12	10.5 dB; 1: 1.85
CW	25.4 dB; 1: 1.11	30.8 dB; 1 :1.06	21.2 dB; 1: 1.19

Shown above are return loss (RL) and VSWR measurements performed on the adapted Serebriakova attenuator shown above. Clearly the input match is better than the output match. The output match did not significantly change when the attenuation switch was moved from 4 to 10 dB attenuation or back.

Based upon these values, it might be a better compromise to put a 3 to 4 dB pad on the input and a 10 dB attenuator pad on the output to ensure an output RL of at least 20 dB. Some might argue that the output RL should be higher. Perhaps, but the match is pretty good for such a simple circuit. Let's put it in perspective; a commercial signal generator that sells on the Internet for $450.00 U.S dollars was measured by a builder I know in the UK and he found a best case RL of 10 dB ! Jim later sold it and built a homebrew signal generator with a 35 dB return loss at all frequencies.

Fixed attenuator pads provide a good remedy for mismatched ports and I discuss why and how in the next section.

The shielded, adapted, Serebriakova attenuator. When home building your personal version, strive to make the AC connections as short as possible. The above device has nearly 23 dB of variable attenuation at 14 MHz. If you can't build, find, nor afford a precision step attenuator for your QRP workbench, this device may work okay for you.

Impedance Matching, Return Loss and VSWR

We radio folks build, buy and apply lots of gear with a stated *nominal 50 Ω input or output impedance*. In truth, a *pure* 50 Ω impedance occurs rarely and components in an RF system are frequently mismatched. Almost every Ham radio operator matches their antenna impedance to their feed line + radio to maximize transmitter output power — but radio and antenna system matching is often the only case where these Hams match their gear.

In contrast, we experimenters, tirelessly match our 50 ohm RF system components — this work flow avails our **modus operandi** on the bench**.** And so, we builders match the input and/or output ports of *all* our RF stages: signal generators, filters, splitters, antennas and so on. You can easily measure the impedance match of your RF components with a basic device based upon a Wheatstone bridge; the return loss bridge. First, let's discuss matching a little more:

On the RF Workbench, we talk about return loss, reflection coefficient and VSWR to quantify impedance matching. I only consider return loss and VSWR on this web site.

When 2 system components are impedance matched, maximal power transfers from 1 device to the other. If the impedances are different, RF power is reflected back to the signal source. This reduces the amount of power delivered to the load. Transmitted and reflected waves moving along a transmission line superimpose and cause standing or stationary waves. The greater the impedance mismatch between the 2 components, the larger the amplitude of the standing waves. Mathematical formulas compute how much power is lost due to mismatch. Consider reading a great tutorial on SWR, Return Loss, and Reflection Coefficient linked here by Wenzel Associates.

Return Loss

Return Loss = the difference between the outgoing incident power and the reflected power as a result of the mismatch between the the signal source and its load. Return loss is expressed in dB as a positive number on this web page. The higher the return loss, the better the impedance match. An ideal prefect match would have a RL of infinity; that is, no power is reflected back to the signal source and all of the incident power is delivered to the load. If a circuit has no load (open circuit), the RL is 0 dB — all of the power is reflected back to the signal source.

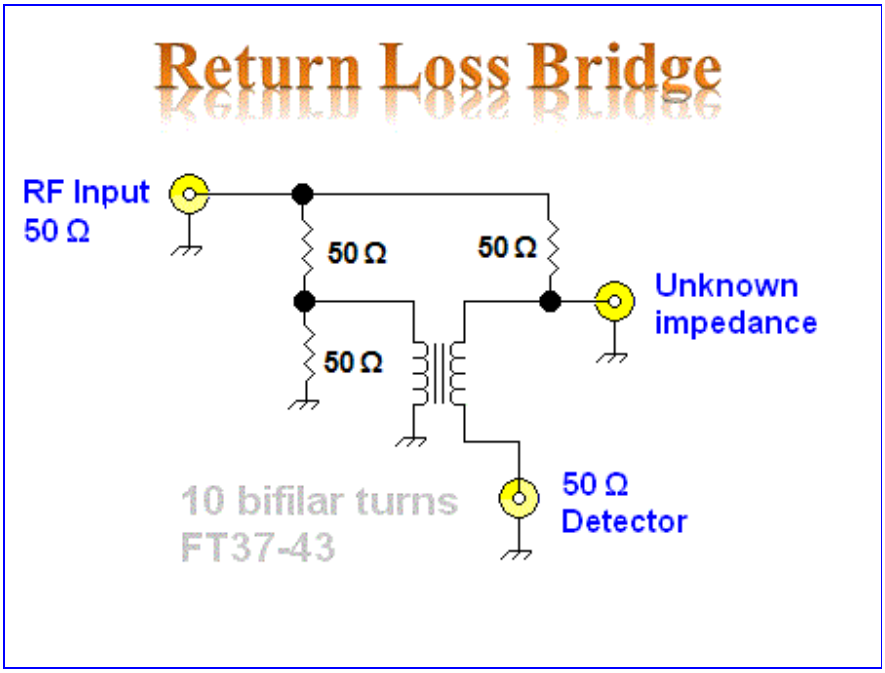
Other terms quantifying return loss are S11 and S22, however S11/S22 are the negative of return loss: RL = 20 dB or S11/S22 = -20 dB. We say S11 as S — one — one and S22 as S — two — two. I discuss these S-numbers, or Scattering Parameters elsewhere.

VSWR

Voltage standing wave ratio is another measure of how well the components in an RF network are impedance matched. Increasing the return loss lowers the VSWR and vice-versa. Most amateur radio enthusiasts are familiar with VSWR and often refer to it as "match" or "SWR". RL and VSWR can be derived mathematically from one other. VSWR = [10^(RL/20) + 1] / [10^(RL/20) - 1]. **Note X ^ Y means X raised to the power of Y therefore 2^3 = 2x2X2 = 8.**

Thus a RL of **10 dB** = 1: 1.92 VSWR and **20 dB** = 1:1.2 VSWR and **30 dB** = 1:1.07 VSWR

In EMRFD, Wes presents a return loss bridge as Figure 7.41. This circuit, shown below is easy to build and use.



The 50 Ω impedance detector may include a spectrum analyzer, power meter, receiver with an attenuator, or a 50 Ω terminated oscilloscope. On my bench, a 50 ohm terminated scope is favored.

Let's measure the return loss of a 27 Ω resistor to learn how. The procedure with a 50 Ω terminated 'scope follows:

- Connect a 50 Ω output impedance signal generator to the bridge RF input port with 50 ohm coax;
- Connect a 50 ohm terminated oscilloscope to the detector port via 50 ohm coaxial cable;
- Record the peak-to-peak (open circuit) voltage with no load on the end of a short coax cable connected to the unknown Impedance port;
- Record the peak-to-peak voltage with "unknown" coaxial cable terminated with the 27 Ω resistor
- Calculate the power difference in dBm between these 2 peak-to-peak voltages.

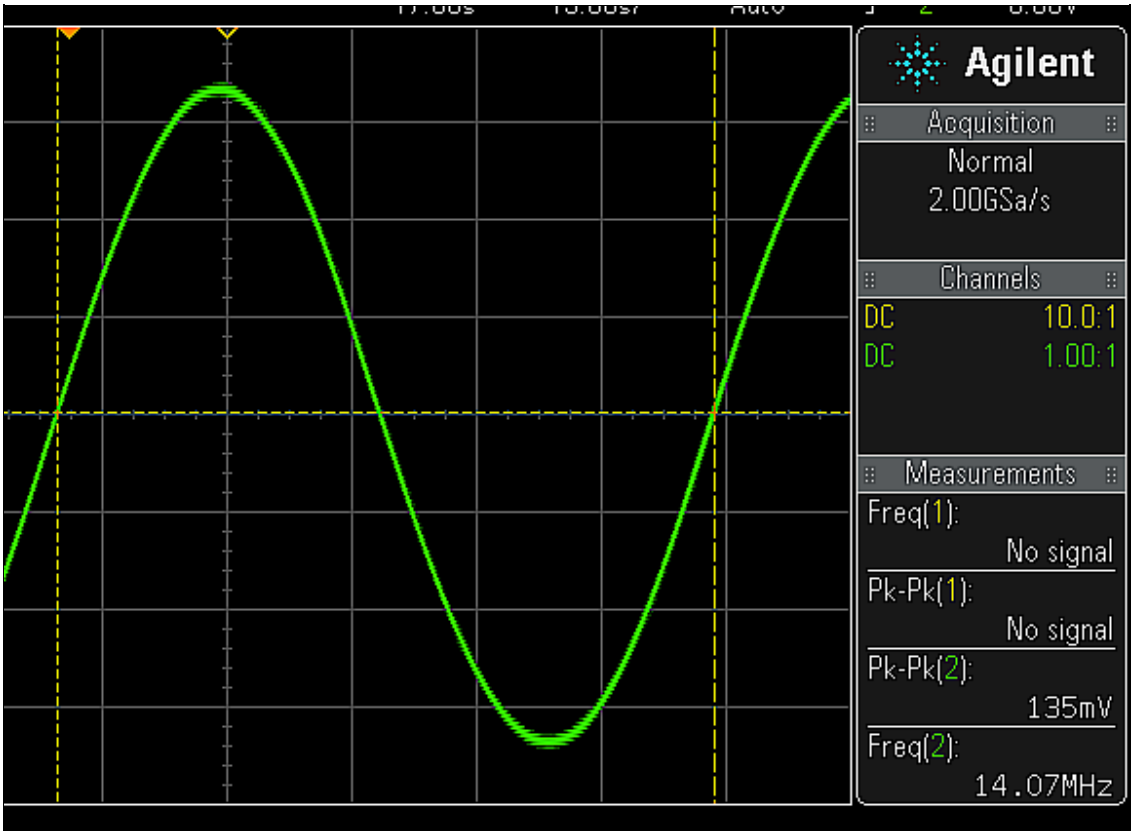
Return loss = the difference in dB between these 2 values calculated by hand or with software. Please refer to the RF Workbench Page 1 for information how to calculate power. I wrote a JavaScript Applet that take these 2 peak-to-peak voltages and calculates RL and VSWR; its labeled **K** on this web page.

Before measuring the unknown RL of a circuit, we usually connect a 50 Ω terminator to the unknown impedance port and calculate the best possible return loss: we refer to this value as bridge directivity — the best possible match for that return loss bridge at that test frequency. I keep a permanent 50 Ω terminator + a barrel connector on my bench for this purpose.

Click for a photo of the gear I use for all RL measurement. *Best viewed at full resolution*

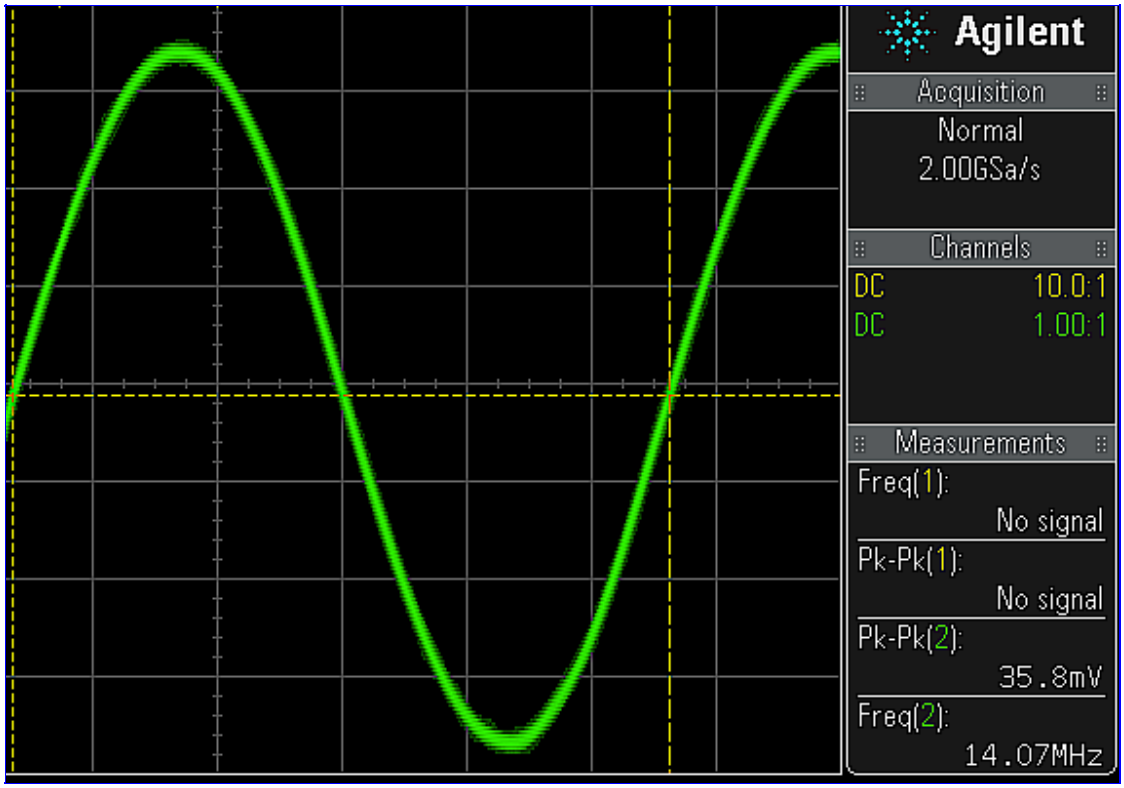
Lets run through the procedure to measure the return loss of a 27 Ω resistor again, but with added photographs and 'scope captures. I tested at 14.070 MHz.

Above — We'll measure the return loss of this device; a 27 Ω resistor soldered on a BNC connector. We call this a resistive terminator and I keep a small collection of 27 - 100 Ω terminators on-hand for calibration purposes.

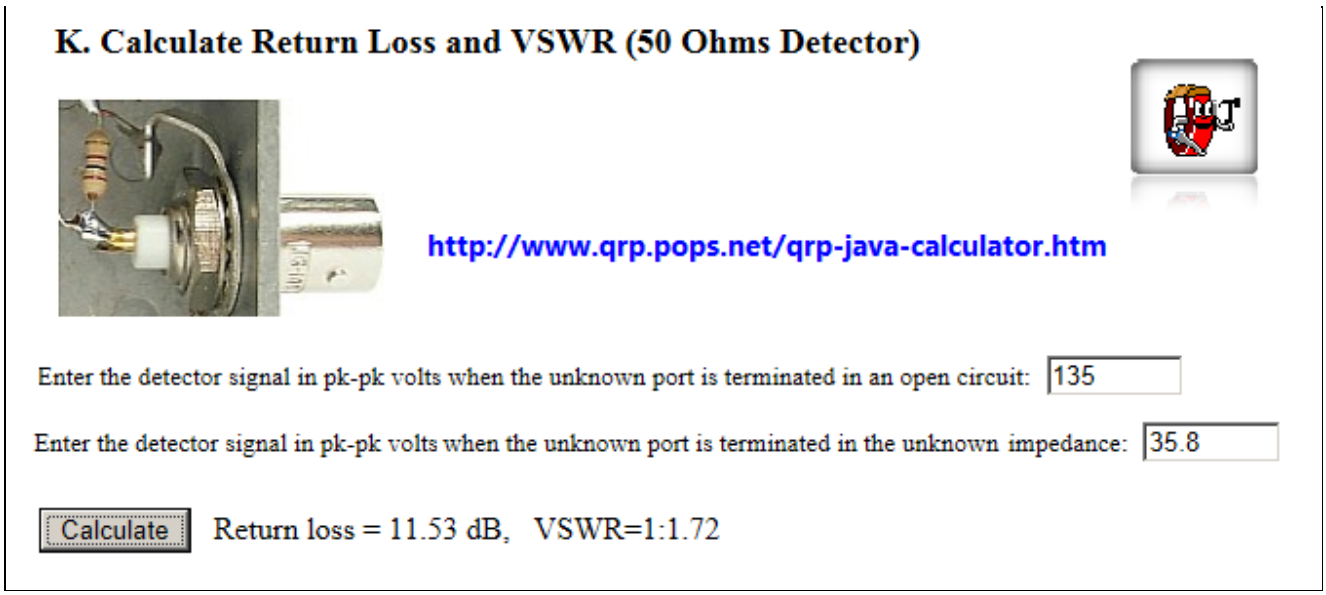


Shown above — The peak-peak voltage with a 14.07 MHz oscillator connected to the RF port; a 50 Ω terminated scope connected to the detector port; and a 20 cm — unterminated — 50 Ω cable connected to the unknown impedance port. **The open circuit measurement.**

Above — Next, I connected the 27 Ω terminator to the unknown impedance coaxial cable with a through-connector interface.



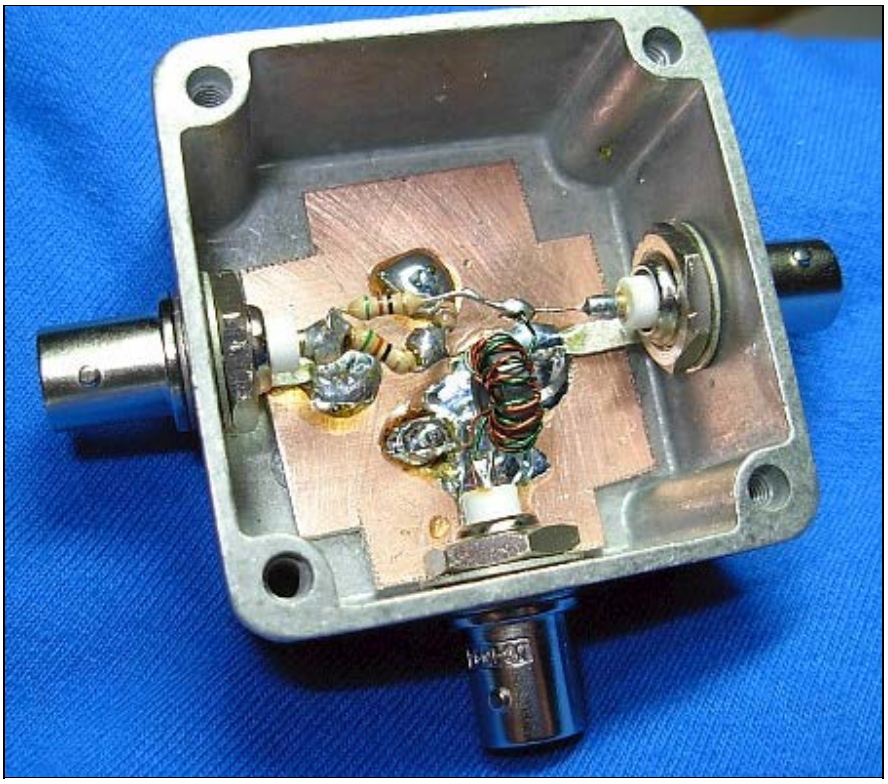
Shown above — The peak-peak voltage with a 14.07 MHz oscillator connected to the RF port; a 50 Ω terminated scope connected to the detector port; and a 27 Ω resistive terminator across the end of the unknown impedance cable**. The reflected signal from a 27 Ω resistor.**



Shown above — Calculating the return loss and VSWR of the 27 Ω resistor with my JavaScript Applet **K**. mV versus volts peak-peak does not matter since we calculate a ratio of power.

Use the return loss measurement procedure depicted above to measure the return loss of any device you choose. If your D.U.T. has 2 ports, terminate the unmeasured port in 50 Ω. I show further RL measurment examples on RF Workbench 3.

I measured the return loss of some commercial gear in my shack and yard and will share 2 examples: 1) An expensive commercial transceiver I borrowed had an input port return loss of 15 dB (a 1.4:1 VSWR) The return loss of 15 dB indicates that the reflected wave power is 15 dB lower in power than the incident wave. 2) With a borrowed commercial bridge, my tuner-matched antenna revealed a return loss of ~60 dB.



A RL bridge from my bench built with 51 Ω 5% tolerance resistors. I show a better RL bridge and some other experiments on RF Workbench 3.

Return Loss and the Attenuator Network (How Do Attenuator Pads Improve Component Matching?)

We routinely employ attenuator pads to increase return loss in a 50 Ω RF environment. For example, let's say you're testing a signal generator and measure a return loss of 6 dB. If you place a 10 dB attenuator pad after the signal generator, the return loss increases to 26 dB. If we used a 6 dB pad instead, the return loss would now = 18 dB. In both cases the return loss is increased by 2x the attenuator pad value. The doubling of return loss occurs because both the incident wave and reflected signals pass through the attenuator pad — that's how attenuator pads

improve matching.

Attenuator pads reduce power, but that is why somebody invented the RF amplifier.

What is the minimally acceptable return loss for a device such as a signal generator? No single answer exists. The minimum return loss depends on the context: are you making precision circuits or just tuning an antenna?

Precision Circuits:

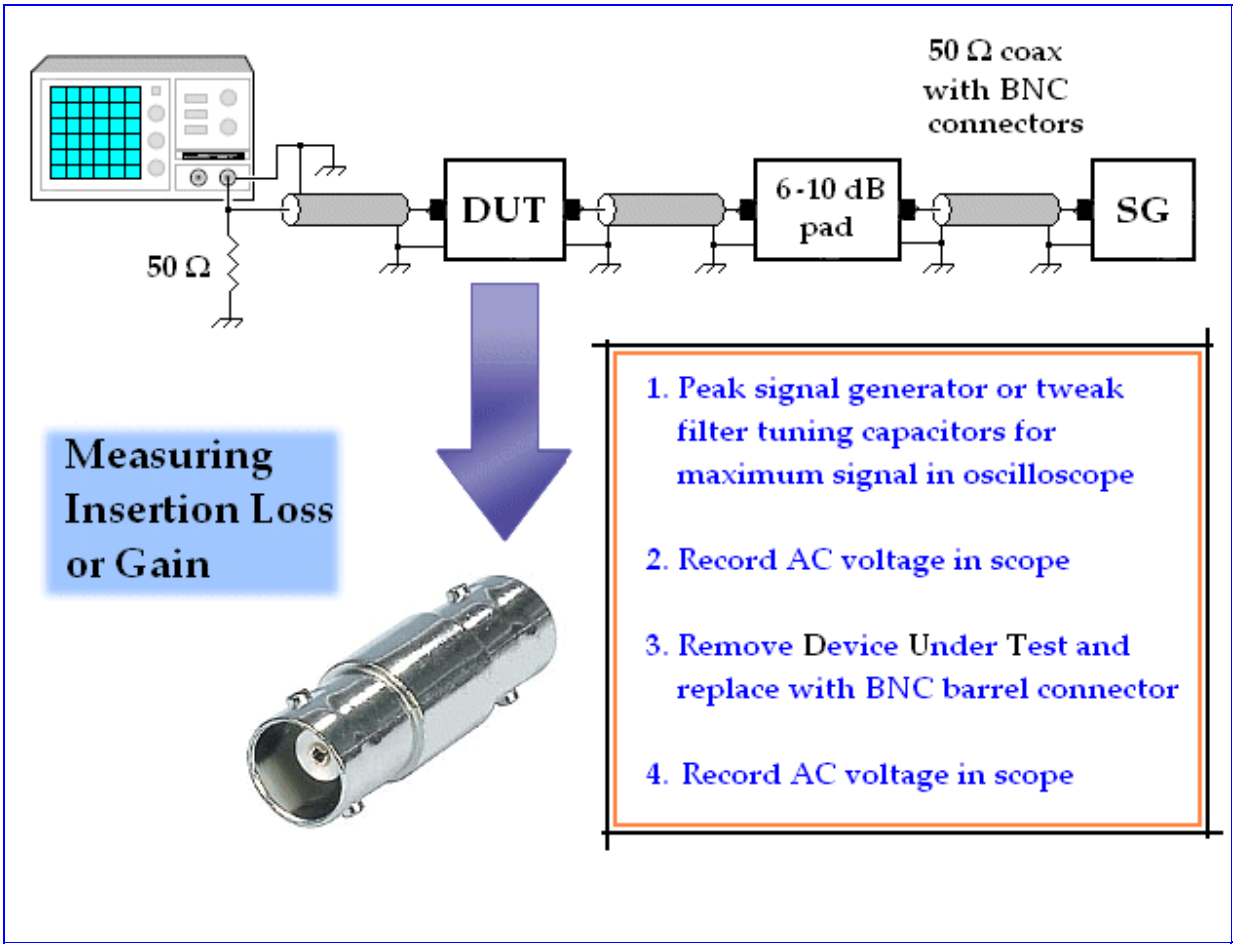
For amateur experimenter bench circuits, aim for a return loss of at least 20 dB. This often means adding an attenuator pad to the ports of your signal generator, amplifier, or other device to get a minimum 20 dB return loss. For an electronic engineer, the minimal return loss is probably higher; maybe 30 dB or so. I have read conflicting opinions about this and for some people — me included — design overkill is normal.

Antenna tuning:

When tuning an antenna for full transmitter output power, the minimal return loss is around 14 dB (a VSWR of 1:1.5). If you measure an antenna system return loss of 14 dB or better, the match is fine. Many Hams will protest a 1:1.5 VSWR and ardently chase a 1:1 VSWR on every frequency with their antenna tuner.

A Method to Measure Insertion Loss or Gain

Often, we want to measure the gain of an amplifier, or the insertion loss of a filter, or attenuator pad. I show how to do this with a 50 Ω terminated scope:



The circuit starts with a signal generator set to the frequency of interest. I show an attenuator pad in this diagram to stress that the signal generator output port must have a return loss >= 20 dB.

Connect the input of the 50 Ω **D**evice **U**nder **T**est to the generator output via 50 Ω coax

Connect the 50 Ω output of the D.U.T. to your 50 Ω terminated oscilloscope

Turn on the signal generator and if needed, peak the signal; In the case of a low-pass filter, the signal generator frequency control is tweaked to give a maximum pk-pk voltage in your 'scope. When evaluating a band-pass filter, tweak the filter trimmer capacitors for the maximum signal

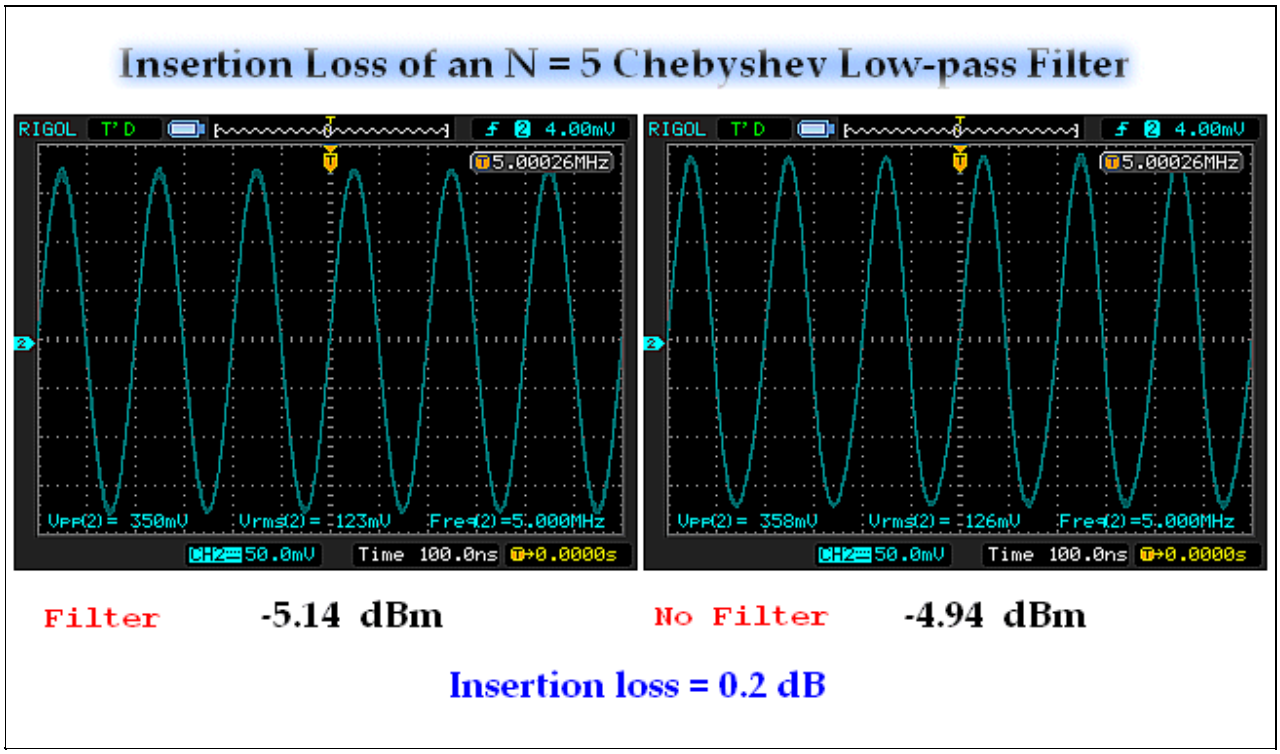
at the desired center frequency. Signal peaking ensures that losses caused by the filters are not caused by the filter mistuning, or in the case of the low-pass filter, to allow for cutoff frequency deviation caused by component value variations. It may be necessary to increase the signal generator amplitude to view a good quality signal in your 'scope.

Record the peak-to-peak voltage.

Remove the DUT and replace it with a BNC through-connector and record this peak-to-peak voltage.

Calculate the power in dBm of the 2 recorded voltages — their difference equals the insertion loss or gain in dB. I wrote a JavaScript Applet to do this. Click and scroll to **H**

This awesome measurement technique controls the input and output impedance and uses the same coaxial cables with and without the D.U.T. for accuracy. Some builders might choose to terminate the D.U.T. with a 50 Ω resistor and measure with a 10X scope. The capacitance of the probe may alter measurement in some cases. As always, choose your measurement technique based upon whatever gear you own and how exacting your standards are.



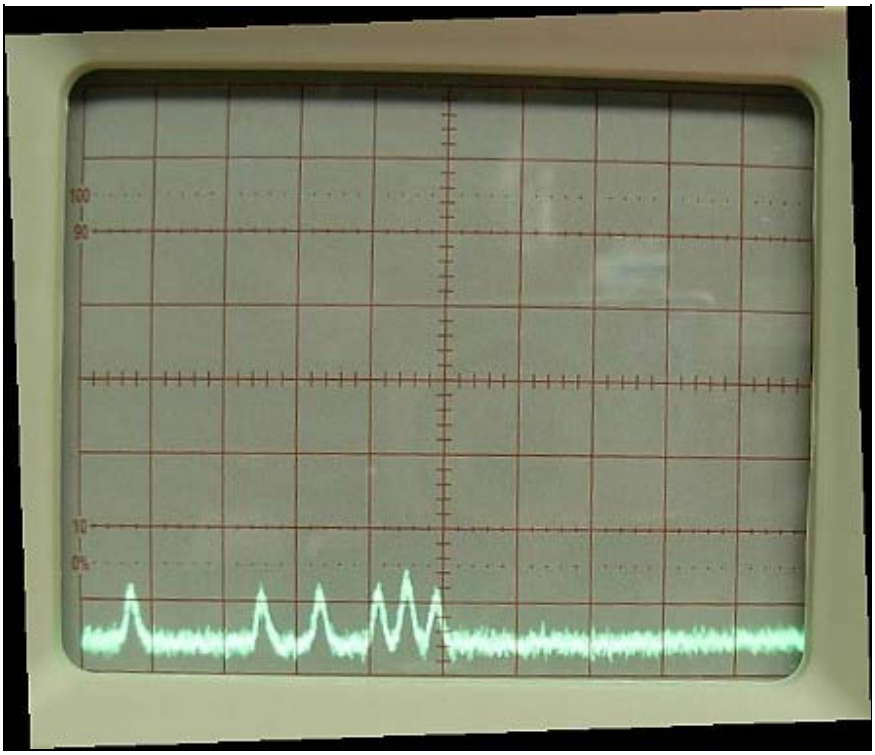
Spectrum Analyzers - Comments from the Workbench

Electronics professionals ruminate that spectrum analyzers are uncommon because experimenters perceive them as esoteric and difficult. My own opinion differs. Spectrum analyzers are relatively uncommon because of one reason - cost. I have watched prices on sites like eBay with amazement. The ads go something like this: 1.5 GHz spectrum analyzer for sale. Built in 1982. Ships in 2 pieces weighing over 22 kilograms. Minimum bid $1850.00. *And...sorry, I live in Florida, U.S.A. and in all likelihood, shipping these 2 heavy pieces is going to cost you a fortune. In the attached ad photos you can see lots of wear and tear, plus some screen burn-in on the display....* Guaranteed to turn on however!

Perhaps I exaggerate or even lampoon the perceived value of old boat anchor spectrum analyzers, but I have bought and sold cars for less money. Be prepared - spectrum analyzers are not cheap. They are however, very cool and open the door into a truly fascinating world. Frequency domain circuit measurement (spectrum analysis) addicts and intrigues. Homebuilding a spectrum analyzer is a serious option, but requires advanced building skills. Click and click for the W7ZOI/K7TAU project.

In recent times, the Rigol DSA-815 spectrum analyzer with tracking generator proved a game-changer to the bloated price of heavy, old and tired gear. Click for a Rigol datasheet. Signalhound also sells spectrum analyzers and tracking generators . A tracking generators *plus* spectrum analyzer allows you to sweep your device under test over a range of frequencies.

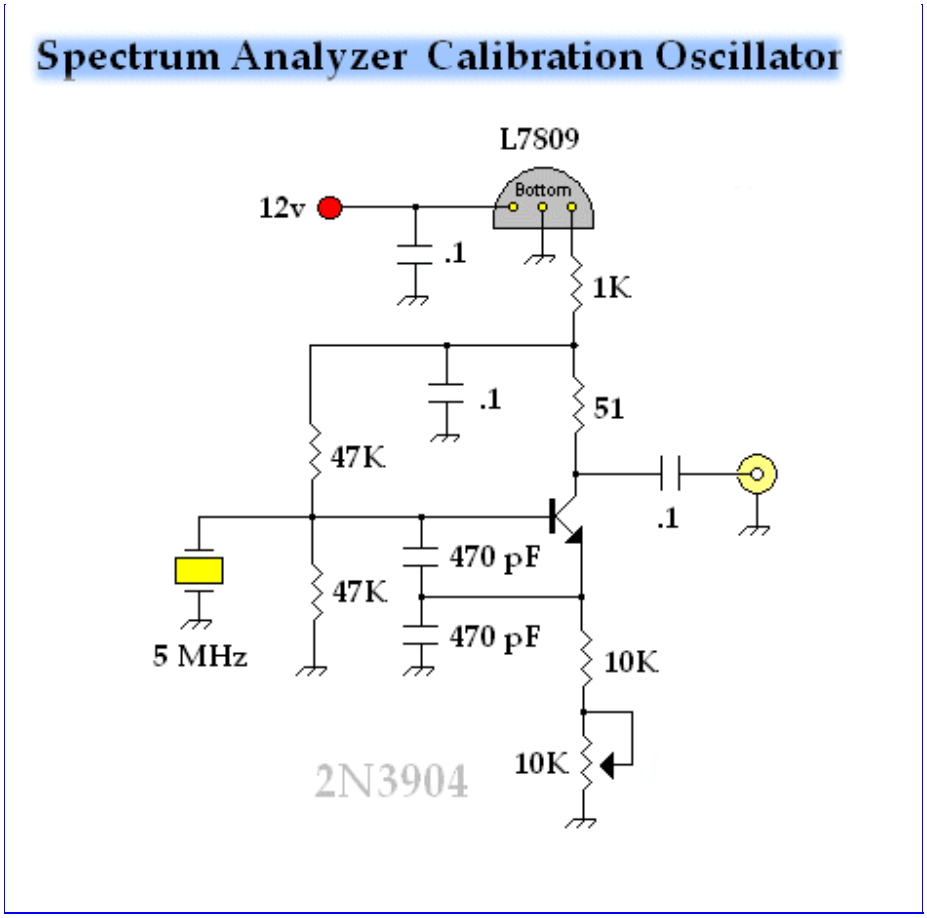
Prior to using a spectrum analyzer, I casually considered shielding stages or placing critical pieces in RF-proof boxes. Quickly I learned that RF in our home and community can and does get into your projects. p>



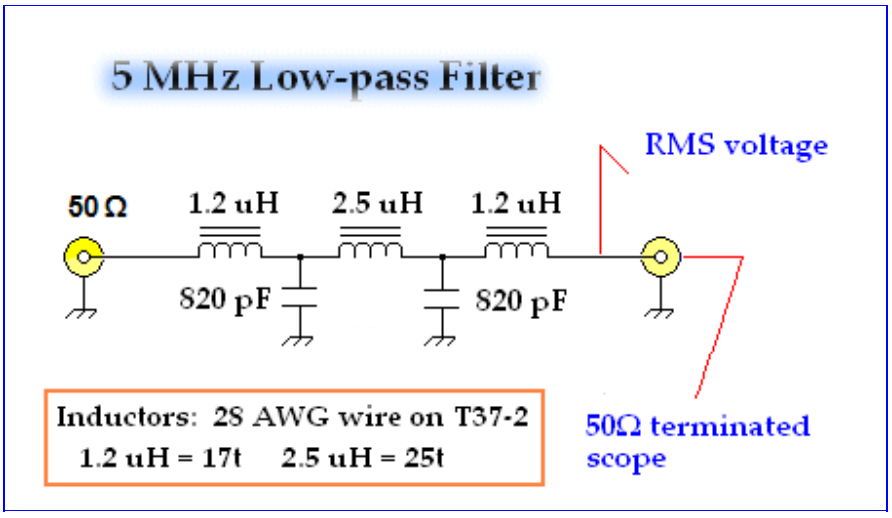
The center frequency of the display = ~150 MHz. The signal spikes appeared and disappeared after 4-9 seconds or so — after a little detective work with my scanner, I learned they were local police and ambulance FM radio conversations. I noticed this interference when I took the lid off a RF-tight band-pass filter — these signals arose in a 28 MHz superhet receiver !! While low in amplitude, experiences like this inform us to watch for lurking RFI.

I found numerous sources of RF in our home with a spectrum analyzer — the clothes washing machine during its spin cycle proved to be the worse RFI generator. RF-tight shielding with SMA, or BNC connectors and DC feed-through capacitors and aggressively decoupling and bypassing DC lines eliminated many of RFI problems during my experiments. I now better appreciate these anti-RFI techniques.

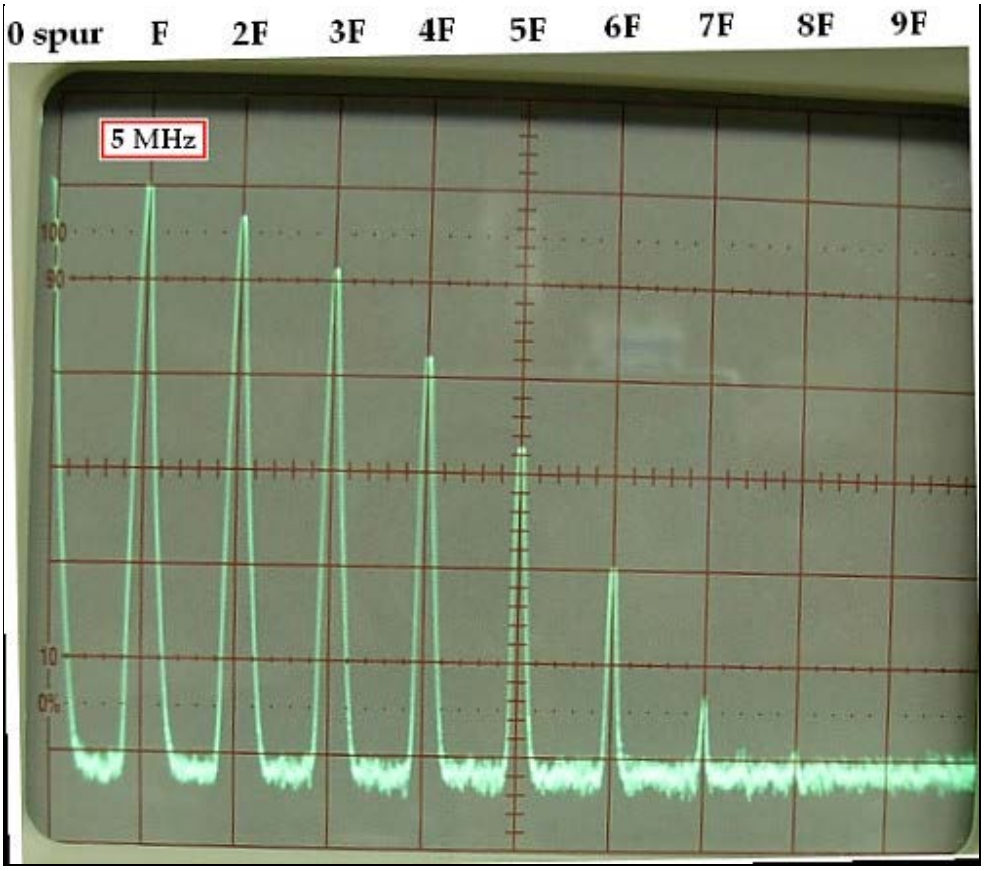
Spectrum Analyzer Calibrator



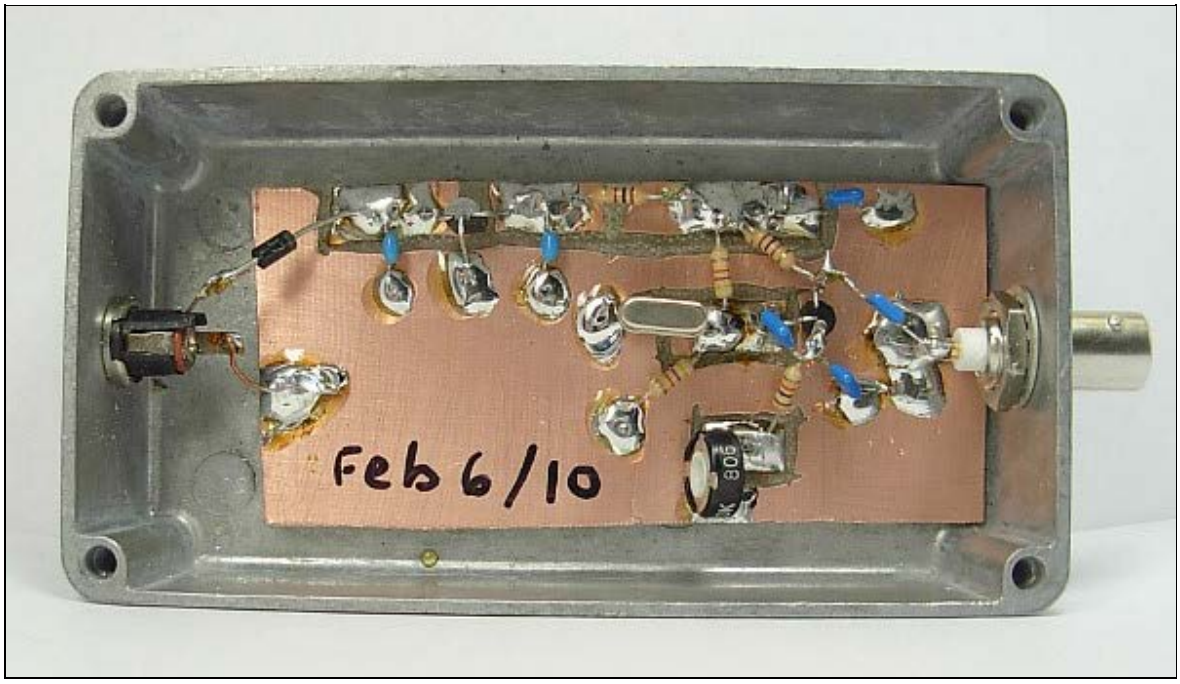
A harmonic rich, spectrum analyzer calibrator designed by Wes, W7ZOI and displayed with his permission. Adjust the 10K potentiometer to provide the output power needed to calibrate your spectrum analyzer. I set mine to -27 dBm. Be careful when connecting signal generators to your spectrum analyzer, since a higher than rated input power may destroy the mixer/front-end of your spectrum analyzer and cost you dearly.



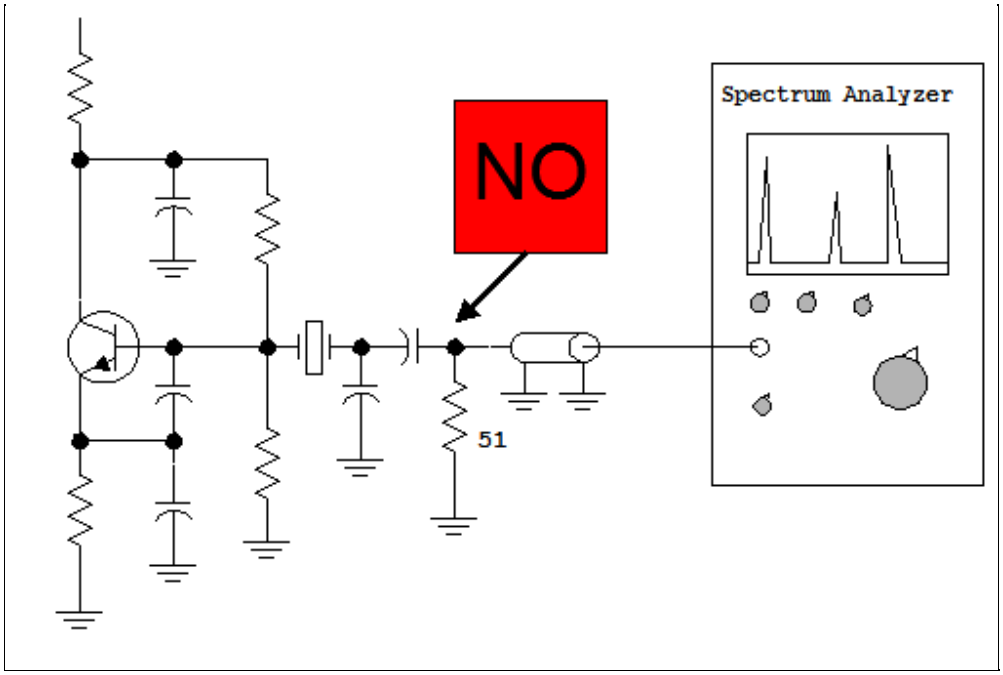
I used this filter to set the -27 dBm power needed to calibrate my spectrum analyzer.



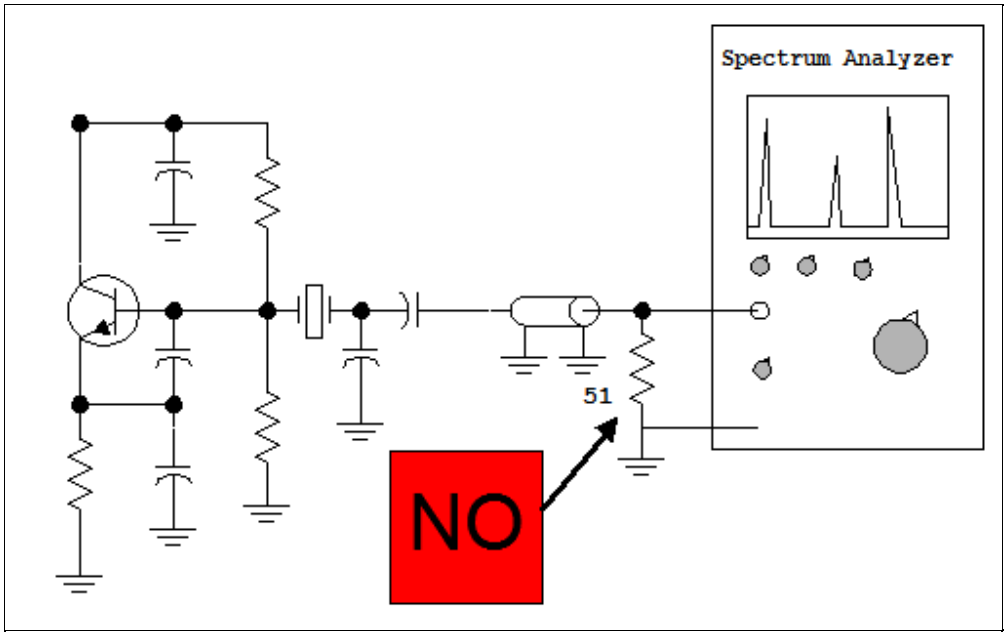
Spectrum analysis of the 5 MHz spectrum analyzer calibrator.



Breadboard of the 5 MHz spectrum analyzer calibrator.



Don't use a "50 ohm" termination when measuring with a 50 Ω impedance spectrum analyzer.



No resistor is required, as the the input impedance of the SA is 50 Ω.

Miscellaneous Photos and Notes

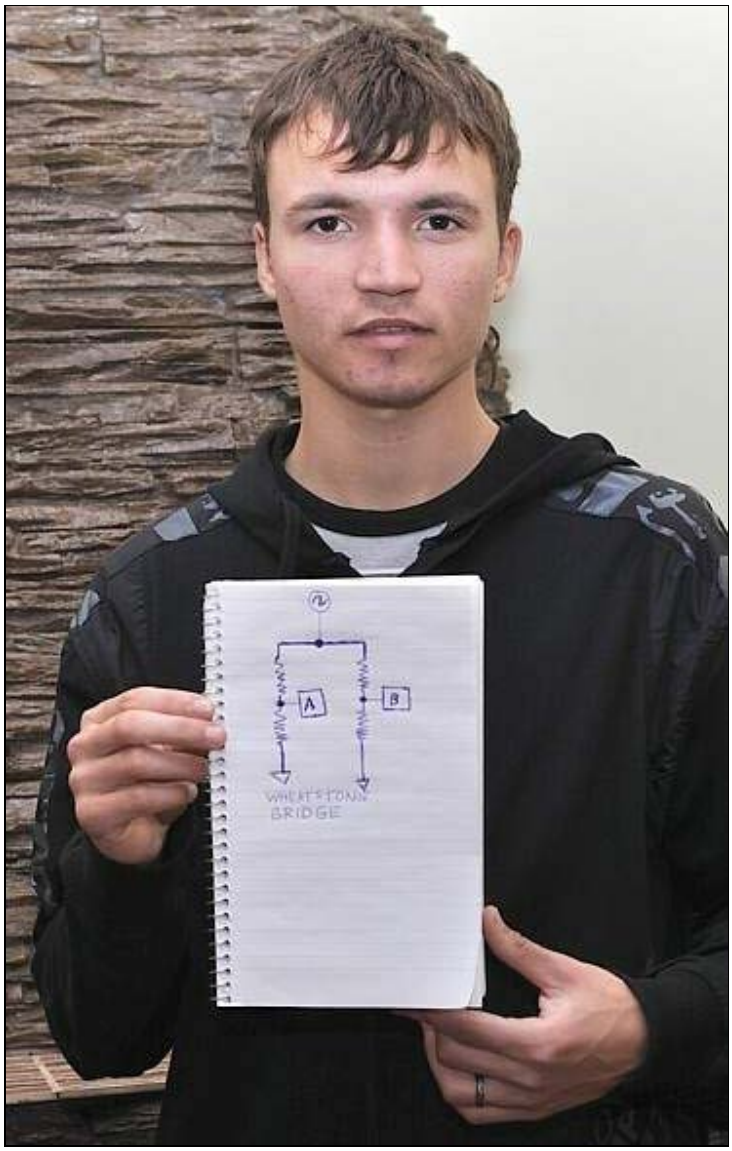
Some of the 50 Ω modules built during the RF Workbench page 1 and 2 experiments

1 3 6 9 30 dB
10 WATT LOAD

15542
HAT-6+
6 dB ATTENUATOR
50 OHM
15542
VAT-6
6 dB ATTENUATOR

Audio Transistor Input Impedance Experiments

Introduction



I examined audio transistor amplifier input impedance during Spring — Summer 2010 and generated enough content for a web page.

On this web page, I explore determining AF amplifier input impedance by using network theory and calculation, plus direct measurement with instruments containing a Wheatstone bridge. This content emphasizes learning through performing bench experiments and I hope it sparks your own experiments and research into impedance measurement test equipment and theory.

Many RF circuits require termination with stages containing a well defined input impedance. Consider, for example, amplifiers that follow L-C low-pass filters, diode ring mixers or crystal filters — a known impedance (usually 50 ohms) must terminate these stages to optimize return loss. A special case is the diode ring product detector; which must be followed by a 50 ohm input impedance audio amplifier. How do we design or assess a small-signal audio amplifier that has a 50 ohm input impedance? This question spawned every experiment on this web page — the content grew and evolved along with my understanding of this topic.

Audio transistor input impedance may be calculated with equations or software, however, doing the math or using or affording these programs might be problematic for some amateur builders. Additionally, component variances such as transistor Beta and different power supply voltages can causes significant differences between the theoretical and the actual input impedance realized. Further, amplifiers, such as a feedback pair that involve combinations of series or shunt feedback can be difficult to analyze accurately using equations during small or large signal analysis. It may be easier for experimenters to just measure and tweak amplifier components on the bench — the focus of this web page!

As an rank amateur, I have much to learn, and by no means am an expert in electronic design. If you see an error on this web site, disagree with my analysis, or have suggestions for improvement, please email them — I am an amateur hobbyist, who earns no money from this site, and who relies on the assistance of others to keep the content as accurate and vibrant as possible.

The topics:

Part 1: Some basic transistor network theory and how to calculate input Z

Part 2: 50 ohm input impedance Wheatstone bridge measurement

Part 3: Measuring unknown impedances

Part 4: Miscellaneous circuits, scans and photographs.

Notes:

Small signal analysis refers to modeling or examining an amplifier at a single operating point (its bias point) and applying linear equations which assess the amplifier with no signal applied. In small signal analysis, we assume that the signal is so small that transistor gain, capacitance and other factors are static.

Part 1: Some Basic Transistor Network Theory and How to Calculate Input Z

Some basic network theory plus methods to calculate the input impedance of common emitter and common base audio amplifiers.

Three basic small-signal transistor parameters include *beta,* emitter resistance *Re*, and bulk resistance *REB*.

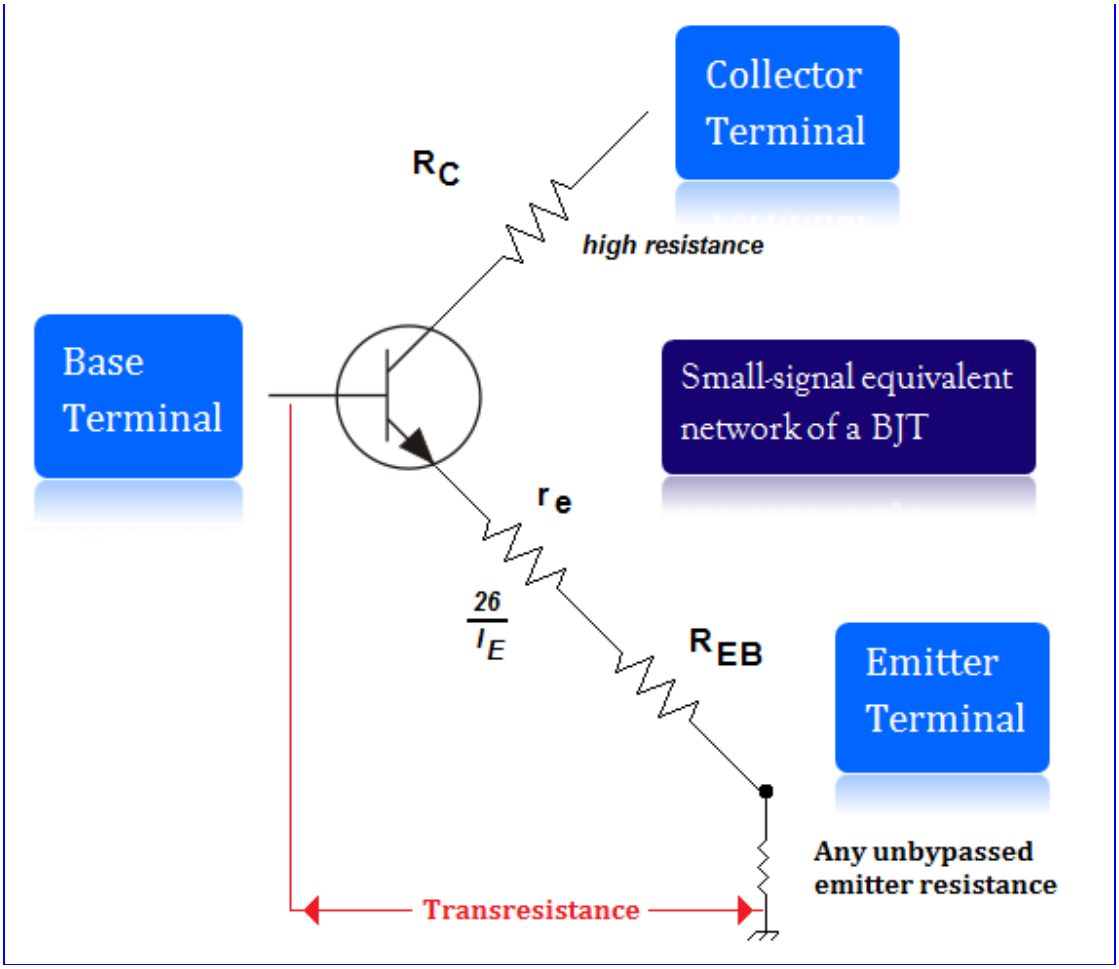
Beta is the term used to designate the current gain of a common emitter circuit — it's the ratio of collector (output) current to base (input) current.

Small-signal emitter resistance; Re = 26 / IE ,or, 26 divided by the emitter current in millamperes. For example, an emitter bias current of 0.52 mA gives a small-signal emitter resistance of 50 ohms, or visa versa . Re is the resistance seen looking into the emitter whether the stage input is the transistor base or emitter terminal. Re is the dynamic resistance of the input junction due to carrier action.

REB represents the bulk resistance of the semiconductor not arising from contact resistance; in other words, it's the DC resistance of the base and emitter leads plus the pn junction. Typically REB = 2 to 6 ohms and is often ignored (your choice) when the current is low — say, for example, < 9 mA for a typical common-emitter voltage amplifier. In large power transistors or for switching operations, the typical REB value may vary. REB, in part, limits the maximal gain of a transistor.

The constant 26 used when calculating the dynamic resistance of a forward-biased PN junction is derived via calculus. Professor Kuhn's website link containing the math.

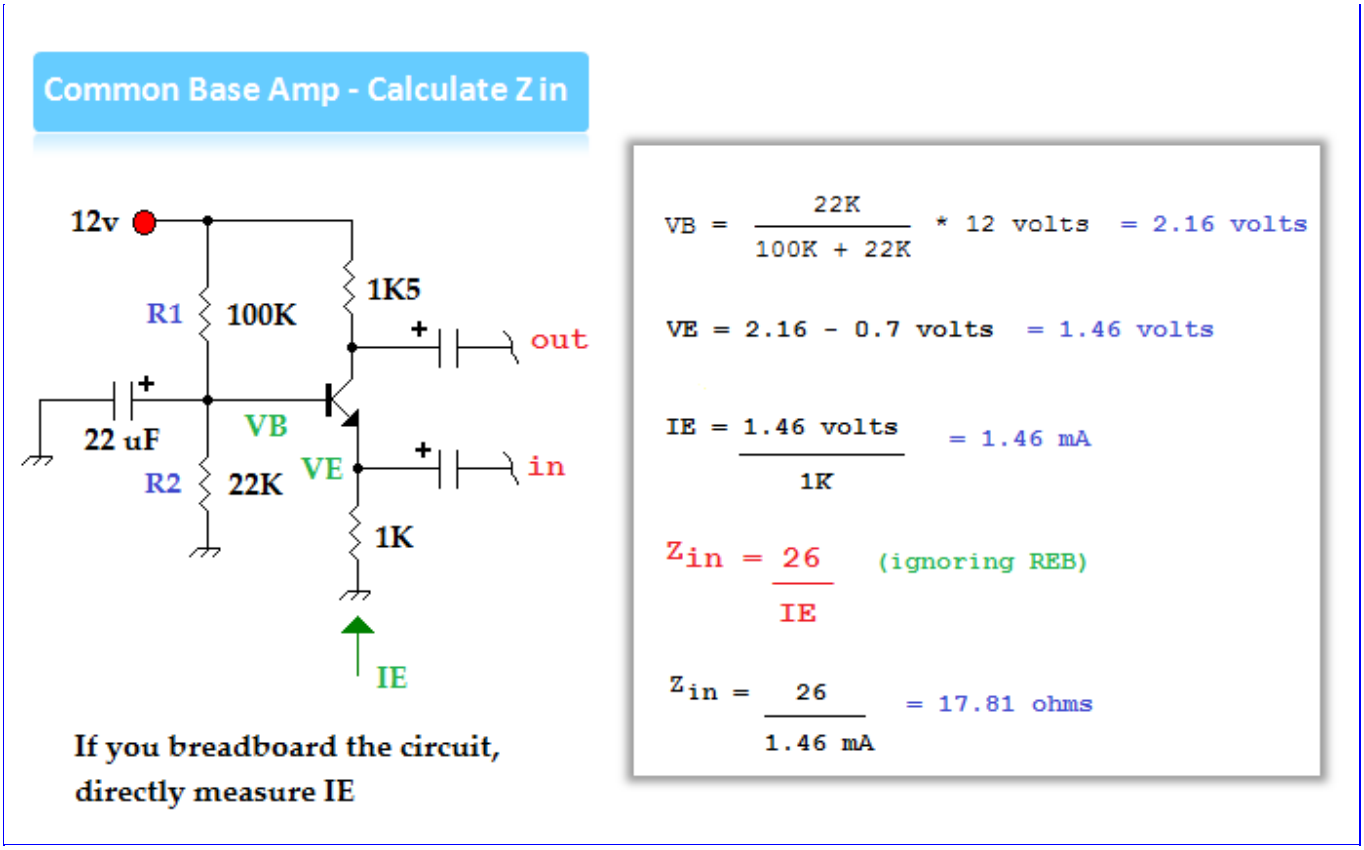
There is also a base spreading resistance generally known as 'rbb' that, in effect exists laterally across the transistor. A simple model puts rbb at about 100 ohms in series with the base and it's one of the causes of finite transistor frequency response. While interesting, rbb isn't discussed further.



Above — the small-signal equivalent network of any transistor. re = 26/IE. Also, re + REB + any unbypassed external resistor may be termed the *Transresistance*, a DC ohmic value representing the total resistance of the emitter. The collector resistance RC is high because of its reverse bias. Collector resistance is not considered when calculating input impedance of simple AF transistor stages.

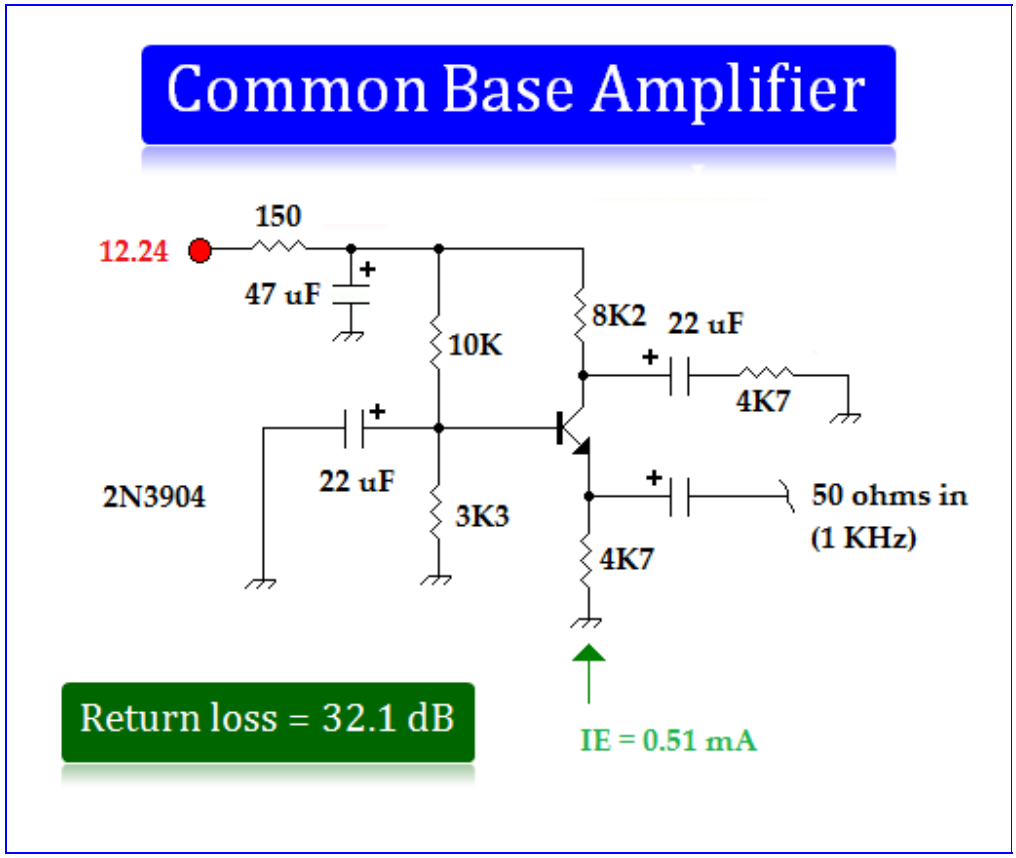
Calculating the input resistance of a common base stage

Calculating the input impedance of a common base amplifier is easy. Input impedance = 26/ emitter current (IE). You can either bench measure or calculate the emitter current using DC analysis. Click for the formula to calculate emitter current . A complete example follows:



Above — An example common base amplifier and its input impedance calculation. In this example, emitter current is calculated using DC analysis. On the bench, it's better to un-ground the 1K emitter resistor and connect your ammeter between this resistor and ground to directly measure IE. REB was ignored and = 0.

Consider the 50 ohm input Z common base amplifier we often use after a diode ring product detector plus diplexer:



Above — A common base amplifier built for a direct conversion receiver in Spring 2010. This amplifier is shown in test setup for bench analysis

— with a DC decoupling network and an AC coupled 4K7 resistive load. The emitter current established using 5% tolerance resistors was 0.51 mA. Therefore, the calculated input Z is 26/0.51 = 51 ohms. The return loss of this amplifier as measured with the *active* 50 ohm Wheatstone bridge device described later on this web page was a spectacular 32.1 dB! If a different power supply voltage or biasing/emitter resistors were used, the IE would change and along with IE, the input impedance and return loss.

This amp illustrates that testing and tweaking AF amplifiers on the bench will garner the best results. If I just copy someone else's design; perhaps with a different DC voltage, or decoupling network and don't adjust the emitter current by tweaking the base biasing or emitter resistor resistors, the input impedance could differ significantly. Whenever you build a common base amplifier, measure its DC current and as necessary, tweak resistors to get the current needed for a perfect impedance match. It is good practice to measure all the DC voltages and emitter current on any amplifier you build — you will learn what is normal, what to expect and perhaps detect errors or parts failure(s).

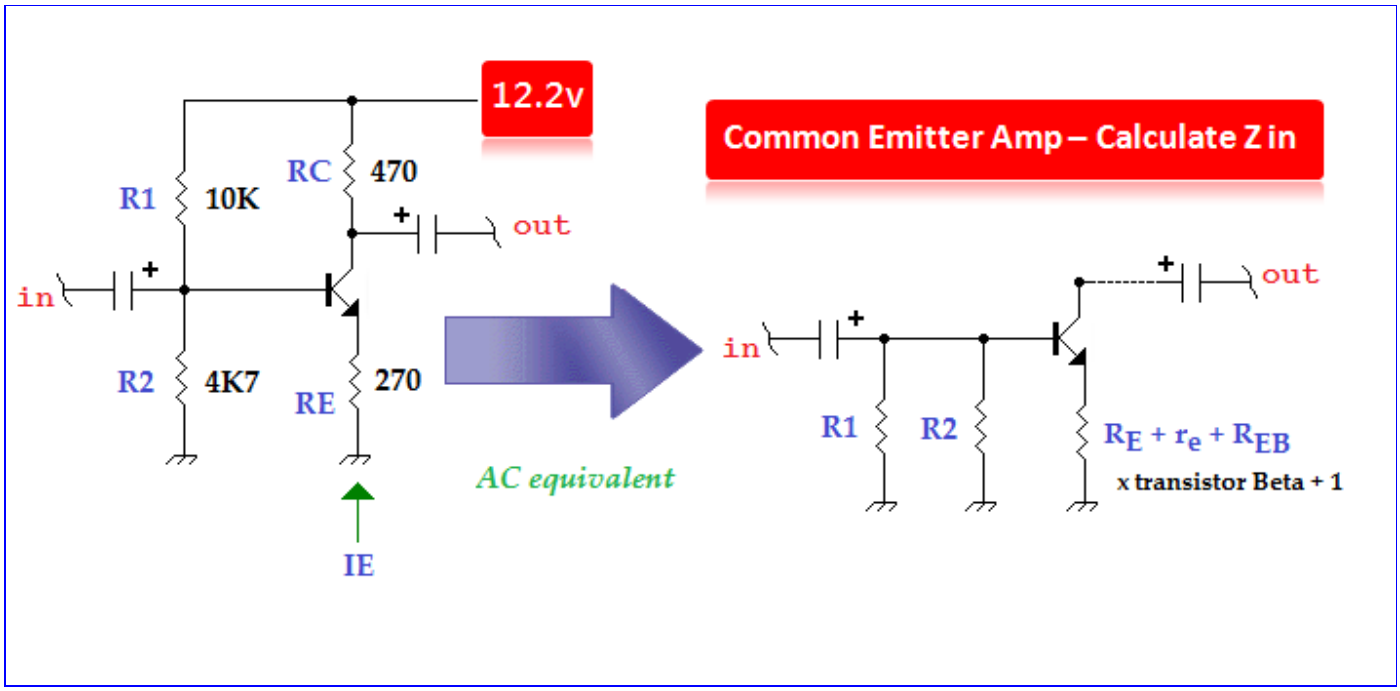
Performing return loss measurement is also a fantastic way to ensure good matching to the 50 ohm impedance diode ring product detector that feeds this amplifier.

Calculating the input resistance of a common emitter stage

Calculating the input impedance of a common emitter amplifier is also straight forward, but not as easy as the common base amplifier.

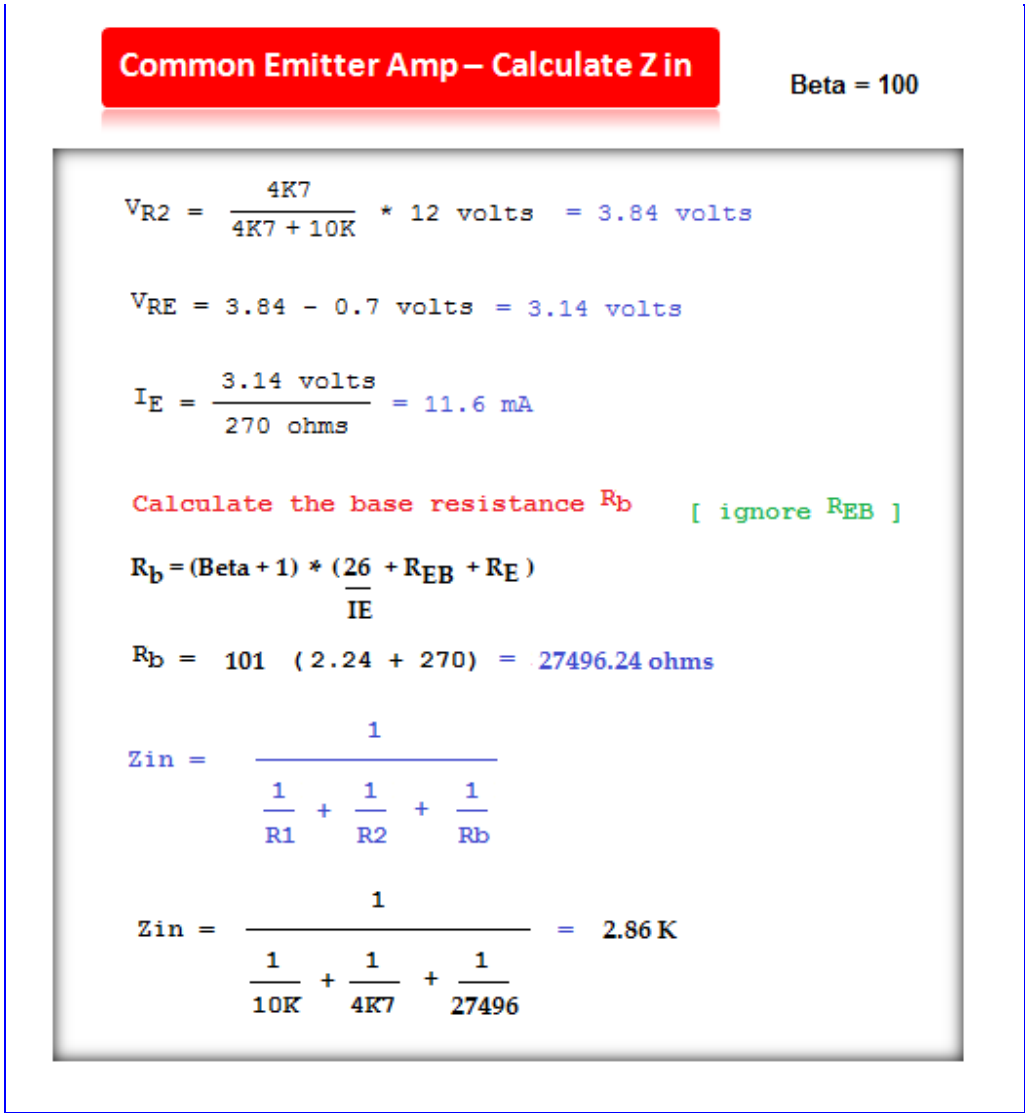
In the **common base amplifier**, the emitter is the input element, therefore the input signal resistance is 26 / IE + REB. Often we ignore REB. If current of the common base amplifier is for example, 2 mA or so, then the 2-6 ohms of REB may be significant as 26 / 2 mA = 13 ohms. REB may be a factor because 2-6 ohms is a significant percentage of the total input resistance.

For a **common emitter amplifier**, the input resistance looking into the base is Beta (26/IE + REB + RE). Again REB is often ignored. We need to include any transistor DC biasing resistors which are also seen by the input signal as it moves through the transistor base. An example follows:



Above — An example common emitter amplifier re-drawn to illustrate how the input resistances combine to provide the AC input impedance. In this case, the 270 ohm emitter resistor RE is un-bypassed. R1, R2 and the components RE, re and REB are in parallel as the DC supply acts as a short to ground for the AC input signal. The components RE, re and REB (if used) must be multiplied by the transistor Beta value (+ 1) since the resistance looking into the base is Beta times that looking into the emitter.

Therefore: Rin = (B+1)*(re + REB + RE'). Normally we ignore REB so practically speaking Rin = (B+1)*(re + RE')



Above — The math for the common emitter circuit shown directly above using DC analysis to calculate the current. On the bench, we just measure the emitter current (no need to calculate it). We assume IE = IC for a common emitter amplifier. REB = 0 (when ignored). If the 270 ohm resistor RE was bypassed with an electrolytic capacitor, the 270 ohm resistance would also = 0; and then Rb = Beta + 1 * (26/IE).

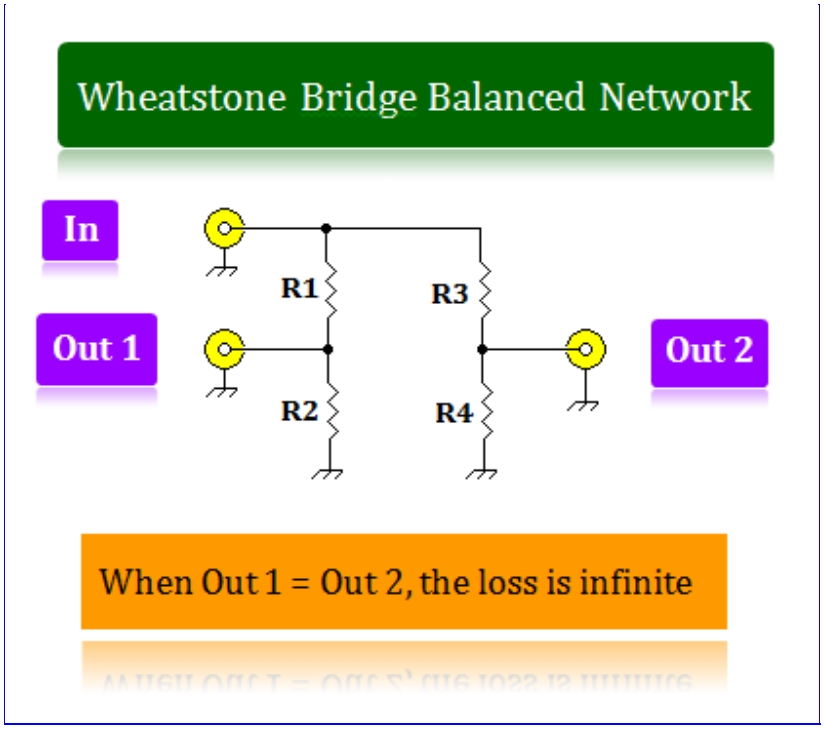
Conclusion

This theory explains how to calculate input impedance in 2 basic transistor AF amplifiers. Consult an electronics text for further explanation. Although the arithmetic is simple, quite frankly, it's a little boring. Let's go to the bench and have some fun. I was quite naive about measuring AF amplifier input impedance; however, my experiments yielded some knowledge and a strong appreciation for the Wheatstone bridge network. Onward...

Part 2: 50 ohm Input Impedance Wheatstone Bridge Measurement

Testing for a NULL or measuring the return loss of AF amplifiers with a 50 ohm input impedance.

Refer to the diagram on the right. Redrawn in a way more familiar to builders, the Wheatstone bridge network is just a pair of voltage dividers in parallel. We measure the difference in AC voltage between the ports labeled Out 1 and Out 2. The bridge is said to be balanced and produce NULL or 0 output when Out 1 and Out 2 are equal in voltage. Another description — when in perfect balance, the signal loss due to mismatch between the output ports is infinite. However, if this balance is disturbed by a mismatch between ports Out 1 and Out 2, an AC voltage appears and the return loss decreases in proportion to the mismatch (within limits and providing your

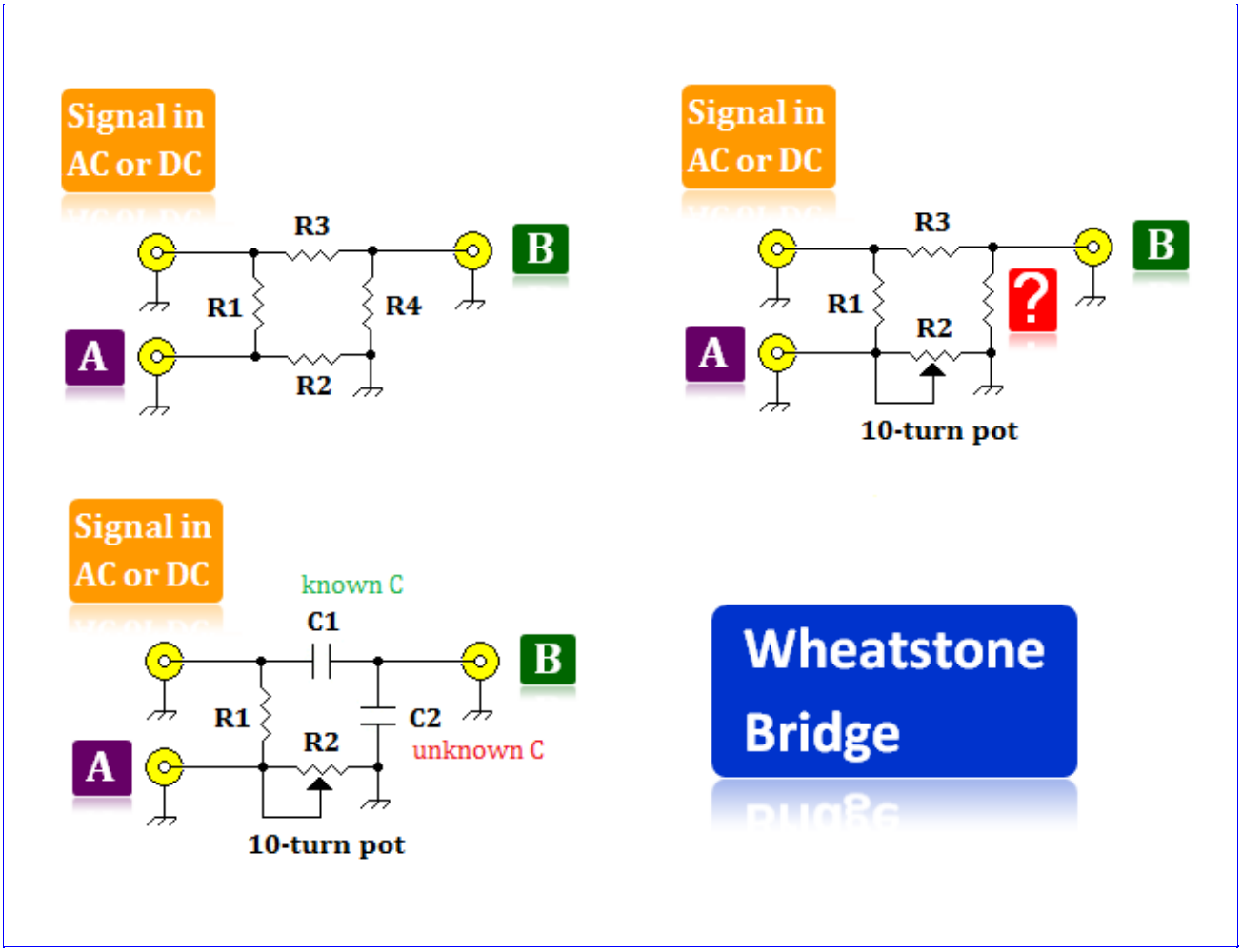


instrument can measure accurately).

Let's focus on some practical bench applications in the new millennia: On your bench, you might employ a Wheatstone bridge network to measure return loss (or VSWR) or to simply to detect a NULL indicating a close impedance match between 2 stages. Specific examples include tuning your feed line and antenna, checking the match between a signal generator and a filter, or measuring an audio amplifier input impedance. In my estimation, the Wheatstone bridge lies among the most important test circuits in the amateur designer's arsenal; worthy of study and experimentation.

Notes:

- The input signal can be AC or DC, but all discussion is confined to an AC signal source
- E96 (1%) metal film resistors were used in all Wheatstone bridges
- All bridges were tested at 1 KHz
- Ensure you do not overdrive your bridge; lest distortion occur! When a bridge is overdriven, you might cancel the fundamental frequency when balancing the bridge, but not the harmonics! Therefore, parasitic harmonics appear in the output that skew the the NULL or return loss values. I learned low pass filtering the amplified bridge output is really important.
- Any distortion in the bridge output means you must reduce your input signal drive level; however, this may reduce the accuracy of the RL measurements. There is no free lunch! You generally want just enough input signal to accurately measure the signal with the bridge at NULL.



Comments from the Workbench

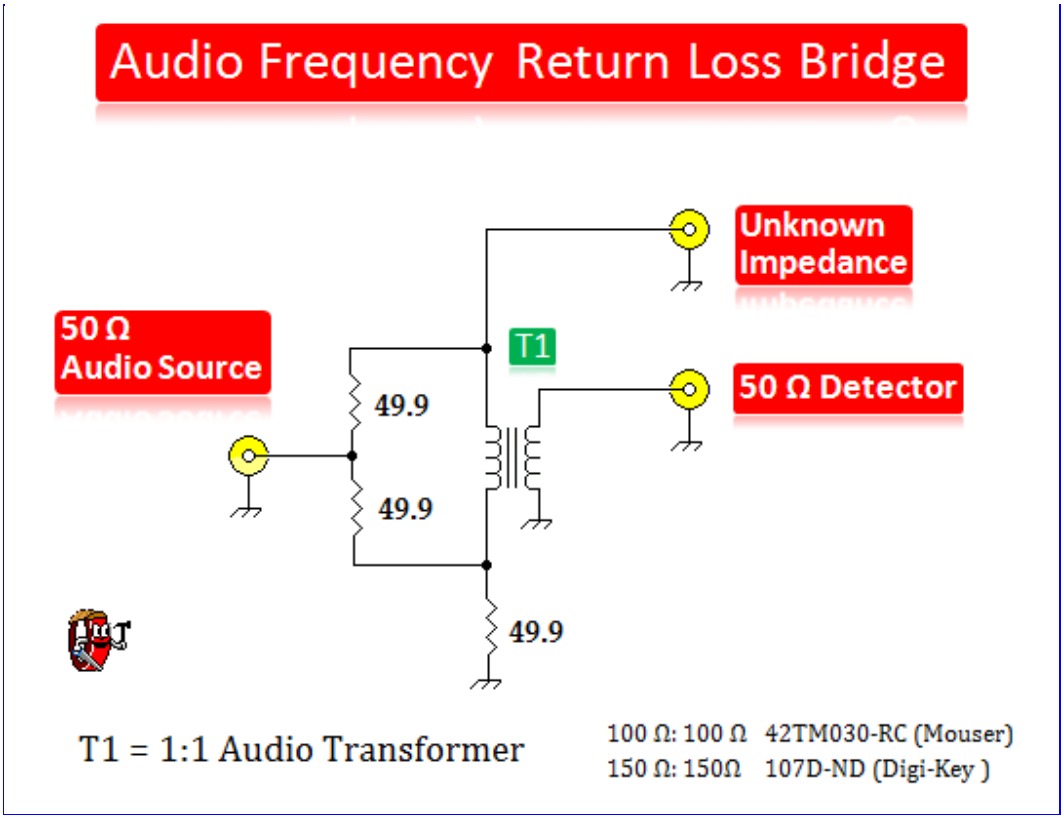
Above — an evolution of the 4 resistor bridge into a device to measure impedance and capacitance. In its classic form, each bridge leg is a resistor voltage divider with a detector connected across ports A and B. If ports A and B have equal voltages and R1 = R2, then R3 and R4 must also be equal; the bridge lies balanced or in a NULL state. If you remove R4 and measure an unknown resistance, the bridge will return to balance after adjusting pot R2 to equal the unknown resistance. In most cases, R2 is calibrated and the impedance is read directly off the potentiometer dial. Bridges can be arranged to measure unknown capacitance, inductance , frequency and other parameters by using precision 1% fixed components, calibrating the 10 turn pot to indicate the desired parameter, or for deriving the unknown value via equations.

Builders of lore used bridges to quantify many values on the bench. Although we have better ways to measure inductance and capacitance today, the Wheatstone bridge is still the *king* when it comes to simple measurement of network impedances; for example, QRP antenna tuners. Some builders use an LED to indicate bridge imbalance.

Building a passive instrument to measure the return loss of a 50 ohm input Z audio preamplifier.

Non-radio folks don't generally understand this — to properly terminate a diode ring mixer, 50 Ω impedance stages are needed. The inspiration driving *all* the experiments on this web page was to design a 50 ohm input impedance common emitter audio preamplifier to follow a diode ring mixer. I could have just used the familiar common-base amplifier popularized by Roy, W7EL, but of course, wouldn't learn anything. Somehow, I became drawn in by curiosity and generated enough content to fill a whole web page.

I decided to try and build some return loss bridges and test them by using known, fixed-value resistors as the unknown impedance. My first bridge, was an AF version of this RF return loss bridge using a junk box 600 ohm, 1:1 audio isolation transformer. It didn't work until I rearranged it as shown in the schematic below.



Above — A simple return loss bridge using an AF transformer and 50 ohm detector. Suitable detectors are described here in the section covering return loss bridges — I used a 50 ohm terminated scope. Using 20 log (peak-peak voltage) to crunch the 50 Ω AC voltage into dB, the bridge was measured at open circuit, plus with various fixed 5% tolerance resistors terminating the Unknown Z port. Using a junk box 600 ohm 1:1 AF transformer, my results initially seemed good, but upon analysis were fraught with error. Note the suggested transformers in the schematic.

Above — The very first AF return loss bridge built. Anchored to the ground plane with resistors, the transformer was a 600 ohm *junk box* special. Although I was able to achieve a deep NULL using a 49.9 ohm resistor, the return loss was 86 dB; not possible. Additionally, other fixed resistors gave return loss values more than 4-5 dB away from the proper value. Likely, my junk box transformer lacked sufficient inductance for 1 KHz. For testing your bridge, use a formula to inform you of what RL value to expect for a given fixed resistor.

Calculate the RL of a resistor in a 50 Ω bridge

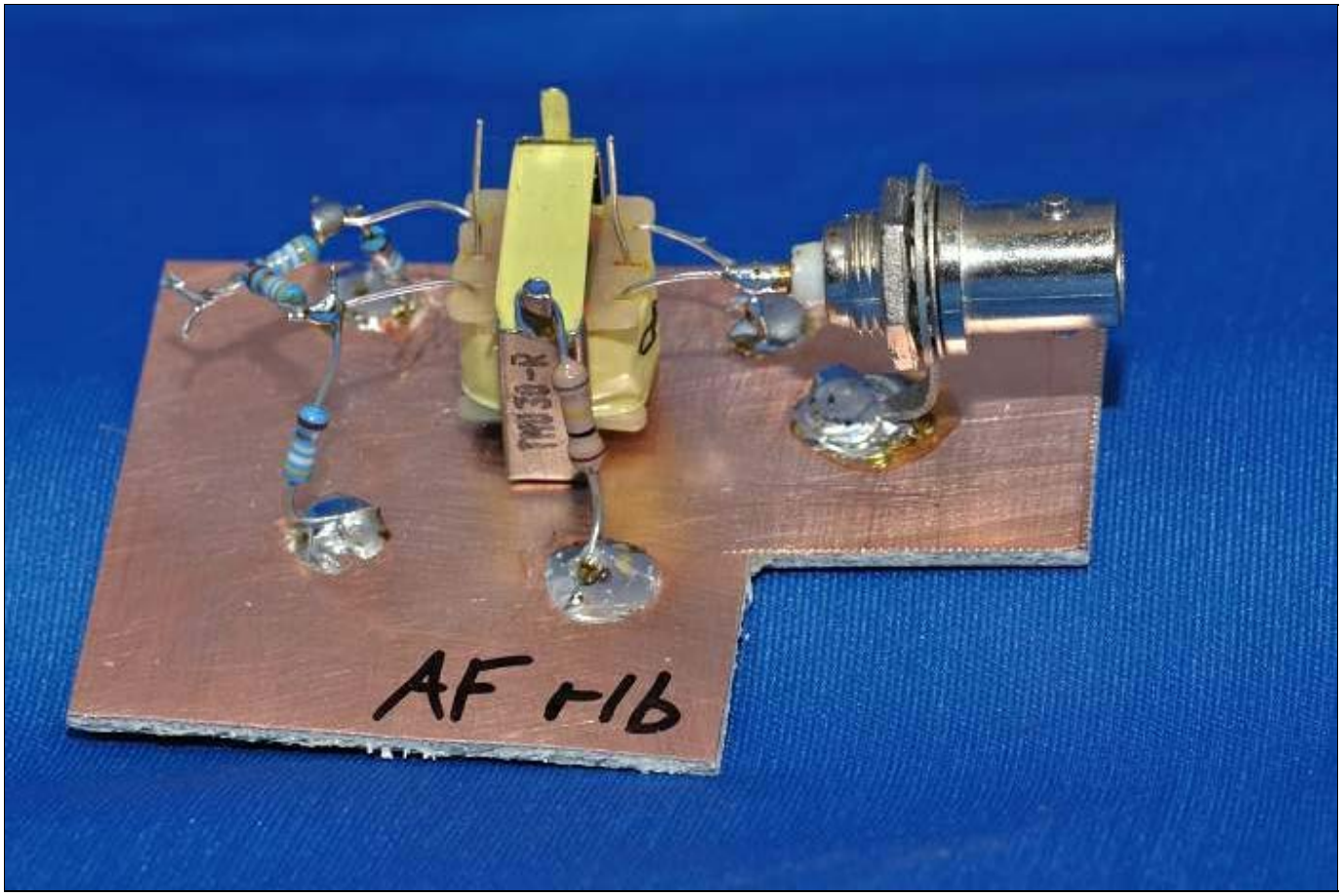
$$V = \frac{(\text{Resistor value} - 50)}{(50 + \text{Resistor value})}$$

$$RL = -20 \log (\text{absolute value of } V)$$

510 ohm resistor:

$$V = (510 - 50) / (50 + 510)$$
$$RL = -20 \log | 0.82 |$$
$$RL = 1.72 \text{ dB}$$

Above — The formula to calculate the expected return loss for a fixed resistor placed in the Unknown Impedance port on a Wheatstone bridge. Click for a table of Return Loss values for some non 50 ohm resistors. Your RL values, will rarely be exact, but should be close to the predicted value. A well functioning bridge should yield a return loss of > 40 dB using a 51 to 47 ohm resistor as the Unknown Impedance.



Above — A second bridge was built after obtaining a 100 Ω : 100 Ω AF transformer from Mouser Electronics. This transformer was ideal (each coil has ~ 1H in inductance!). Bench testing indicated good function. My results are tabled below:

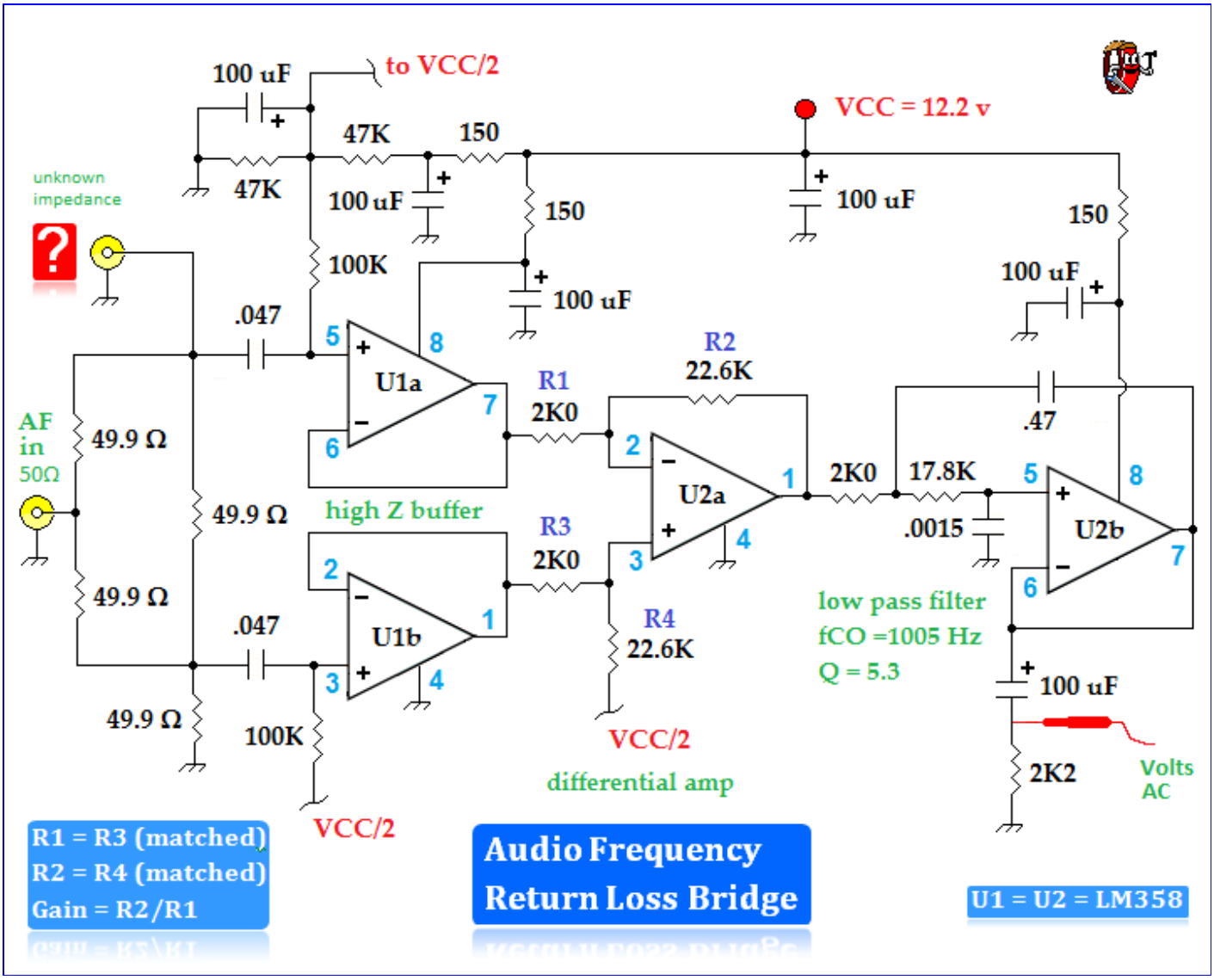
Measurement of R L B	
Resistor Ω	Return Loss (dB)
10	3.35
22	7.47
27	9.74
33	12.39
39	16.23
49.9	56.73
62	16.54
75	12.82
100	9.15
220	3.95
510	1.63

Above — A table of the above 50 ohm Wheatstone bridge return loss measurements. These results are acceptable. The NULL with a 49.9 ohm resistor was incredibly sharp and garnered a RL of 56.73 dB. My AF source was a low noise 1 KHz, 50 ohm output impedance signal generator.

If you do not need return loss, and only require a NULL to indicate a match, a common 600 ohm transformer may work okay for you.

Building an active instrument to measure the return loss of a 50 ohm input Z audio preamplifier.

The results of my early experiments with a passive bridge were encouraging. Noting that most builders would have difficulty obtaining a 100 Ω : 100 Ω AF transformer, a version using op-amps was sought. My first 3 designs did not work properly and I became discouraged. Some guidance from Wes, W7ZOI allowed me to problem solve and experience success.



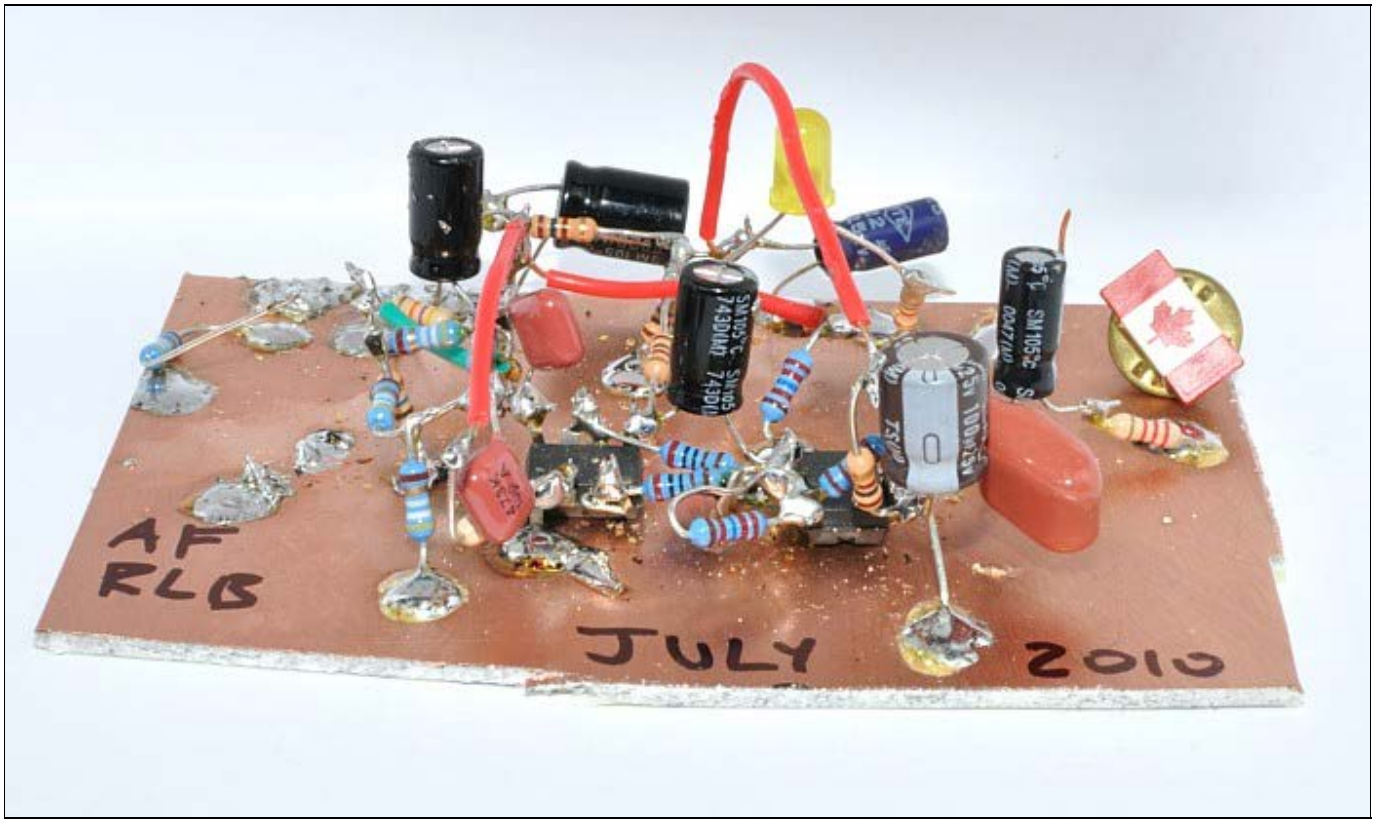
Above — Schematic of my active Wheatstone bridge, amplifier and low-pass filter for measuring the Return Loss of 50 ohm input impedance AF amplifiers. I built 3 copies of the above device; best results occurred when careful layout and planning were employed. Optimal performance occurred when encased in a metal box.

The bridge was built from 1% metal film resistors. 0.047 polyester film capacitors lightly couple the bridge to high impedance op-amp buffers labeled U1a + U1b. My experiments informed me that to minimize loading on the bridge is important. The LM358 is an excellent op-amp choice, but almost any other op-amp could be employed successfully. U2a is the differential amplifier and matching R1 + R3 and R2 + R4 with 1% tolerance resistors is critical; 5% resistors did not work well. The gain is non-critical — feel free to choose reasonable resistor ratios based upon the resistors you have in stock. The differential amp promotes the unfortunate side effect of amplifying both the desired AF source plus any common mode signals. Although common mode suppression is an important consideration when designing instrumentation amps, fortunately, performance is fine. A amplifier topology using a differential amp across the bridge was trialed, but functioned identically to the simpler differential amp shown. Consult a textbook for more information on Instrumentation amps. Much information was gleaned from Professor Ken Kuhn's web site.

The output is low-pass filtered by a single stage Sallen-Key low-pass filter with a peak frequency of 1 kHz and a Q of 5. This filter gain at 5 at 1 kHz is 0.328 at 2 kHz and 0.123 at 3 kHz. Thus, the second harmonic is reduced by a factor of (5/0.328) = 15.2 and the third harmonic is reduced by a factor of 40.6. **Do not omit a low pass filter**. I chose a 1 KHz cutoff, but experimentation indicated a low-pass cut-off frequency as high as 10 KHz may work okay if you plan to use the bridge at frequencies other than 1 KHz.

Power supply decoupling proved important. When less DC low-pass filtering (less than the 150 Ω plus the 100 uF capacitors shown) was employed, some low frequency audio noise appeared in the output.

I measured using a X1 oscilloscope probe on the output of U2b.

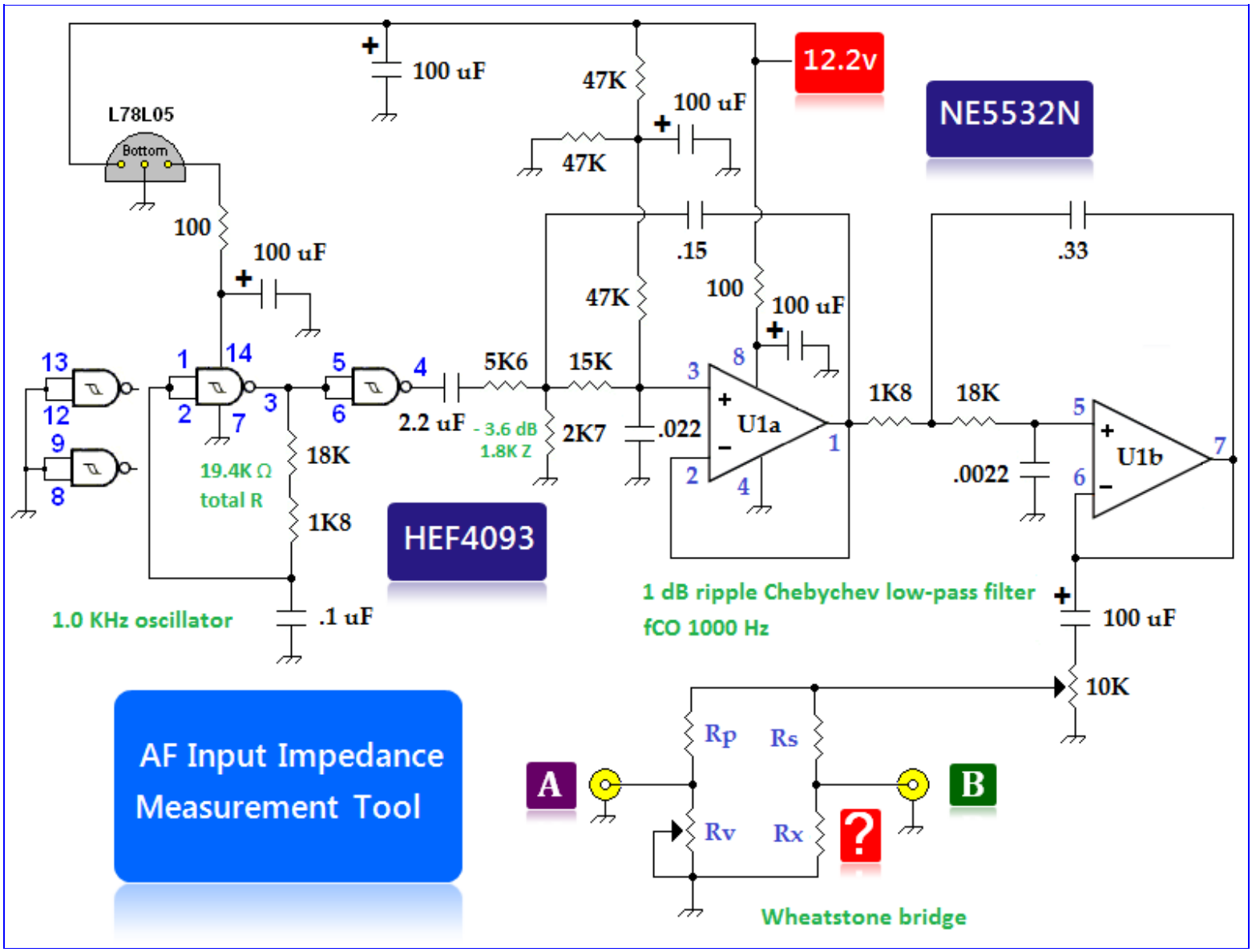


Above — A breadboard of the active Wheatstone bridge schematic located above. When tested with fixed value resistors, the RL @ 49.9 ohms was 55.4 dB and close to predicted value with other test resistors. This instrument will be put in a metal case and become a permanent part of my test equipment arsenal.

Part 3: Measuring Unknown Impedances

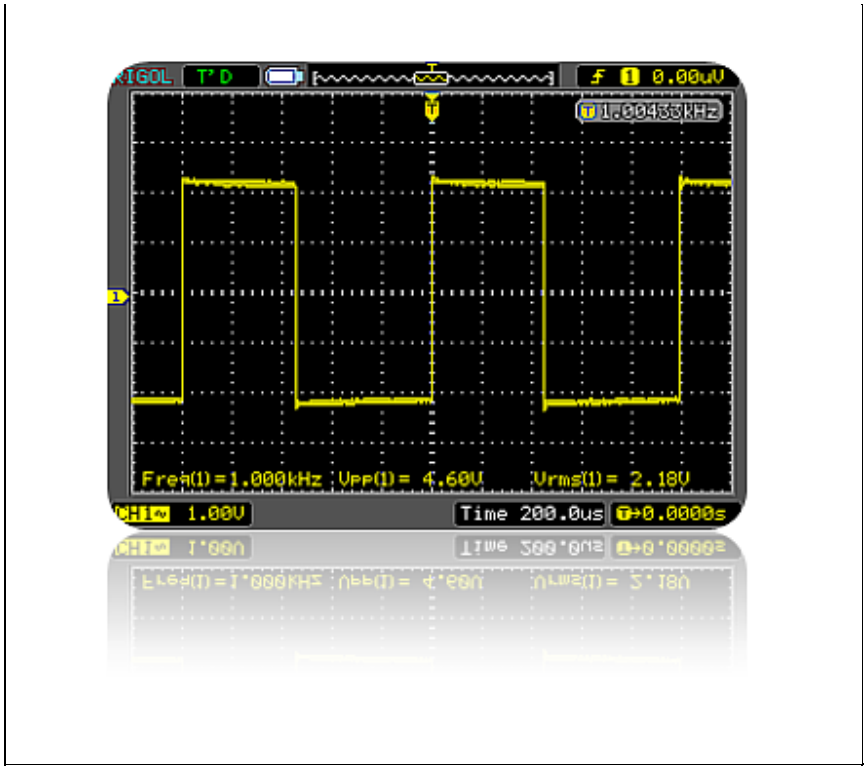
Building an instrument to measure the input impedance of an audio preamplifier using a NULL.

Above — One of several Wheatstone bridge circuits built in the Spring-Summer of 2010. In these bridges, the potentiometer was calibrated and the panel labeled using fixed resistances for calibration. One big challenge is range or resolution; dependent on the bridge resistor values and what impedance you are trying to measure. Greatest accuracy is associated with 5 or 10 turn potentiometers, but these are expensive. Often, I used standard, linear taper pots to save money during my experiments. Over 7 different bridges were built and tested. To save time, I didn't photograph many of my projects from the summer.

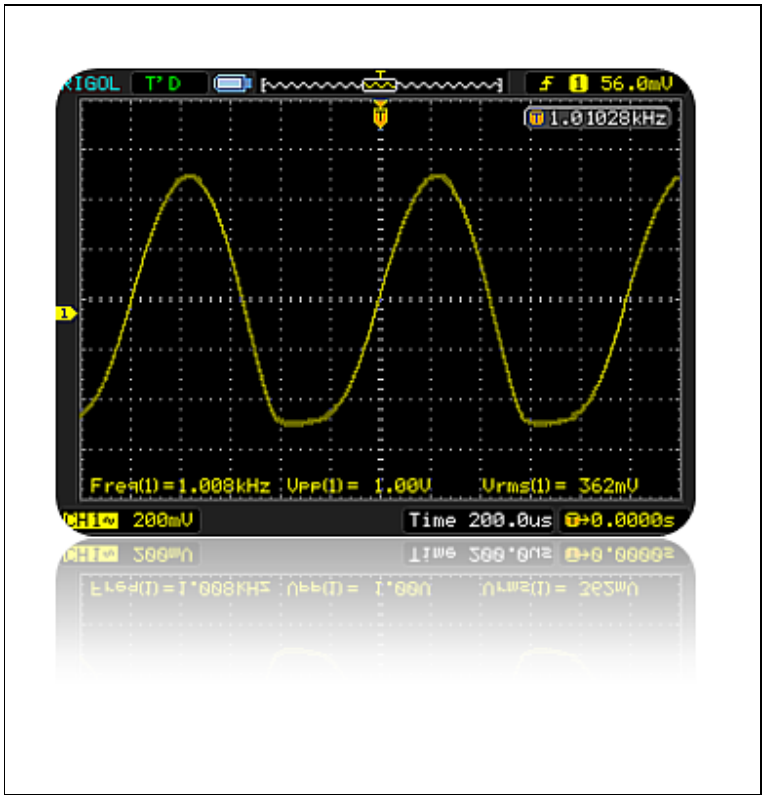


Above — A complete 1 KHz signal generator, low-pass filter and bridge circuit which became the prototype for most of my experiments in this section. Click for a high resolution photo of 1 of the breadboards during construction.

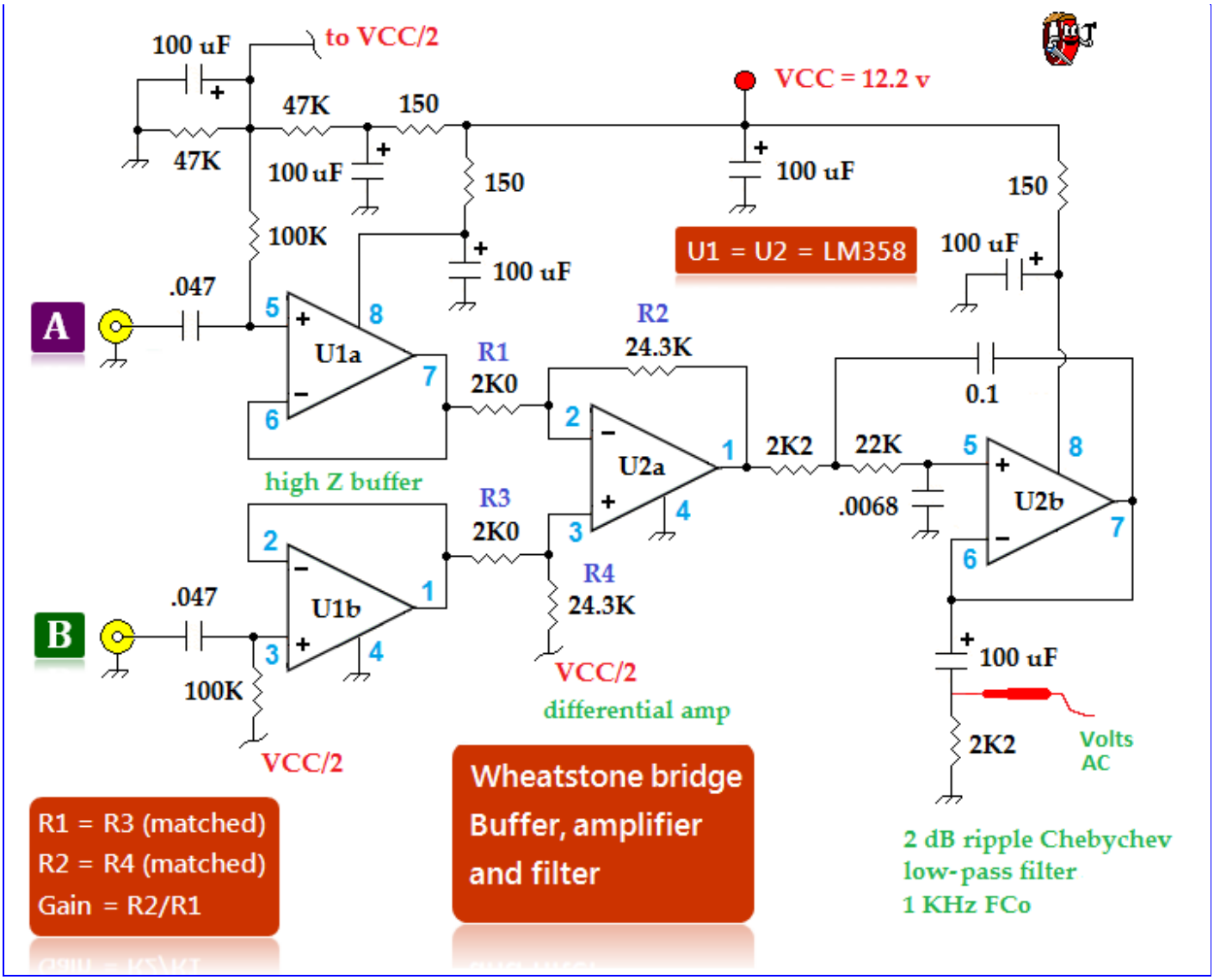
The 1 KHz signal generator is a digital oscillator built with 2 gates from a 4093. This excellent oscillator uses a single R + C network for tuning and requires a voltage regulator for frequency stability. The output signal is attenuated 3.6 dB and low-pass filtered by 4 poles of active filtering. A 10K pot controls the drive into the bridge circuit. The bridge outputs are labeled **A** and **B** and require buffering, amplification and low-pass filtering similar to the active bridge shown earlier. These functions and some comments on the bridge resistors come later.



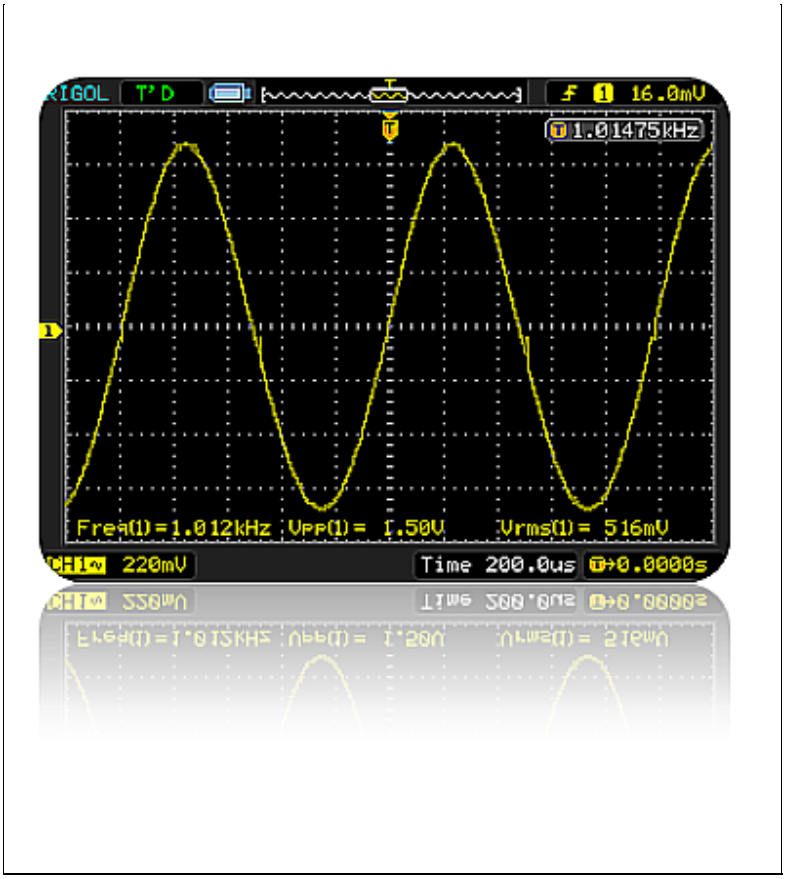
Above — The oscilloscope waveform from the digital 1 KHz oscillator. Digital clocks fascinate me and this was an untried design. Initially a CMOS 555 timer was considered, however, I own many 4000 series NAND Schmitt triggers and pressed 1 into service. Another good choice might be the 74HC132. The first NAND gate (inverter) contributes 180 degrees of phase shift, while the RC low-pass filter tank circuit digitally shifts the AC the other 180 degrees. Output noise is filtered by both the low-pass filter and the input Schmitt trigger dead band or hysteresis. The result is a fairly crisp square wave that may rival the 555.



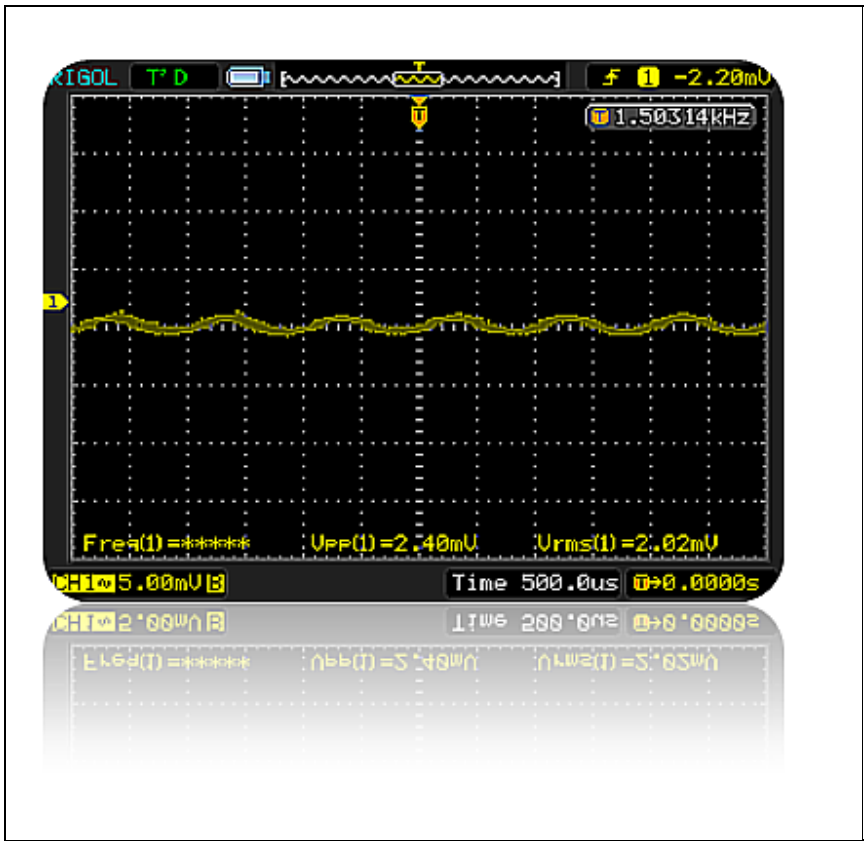
Above — The output of the 1 KHz low-pass filter. A sine wave is desirable, but not critical; suppression of energy in the range of 5-10 KHz, informed the filter design goal. At 8000 Hz, the attenuation is > 80 dB. All good.



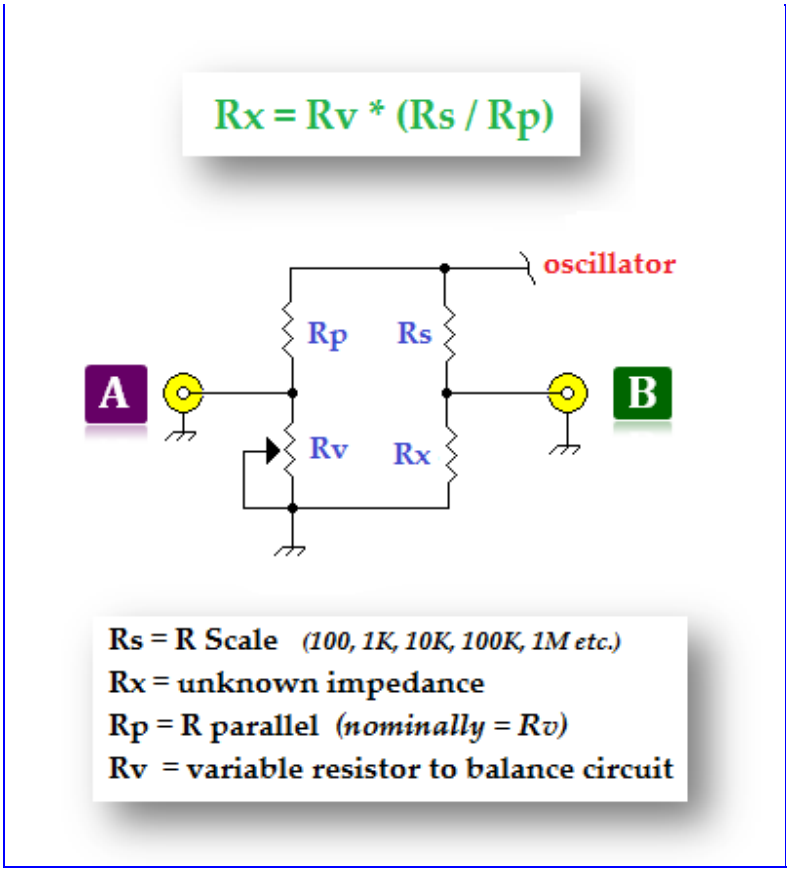
Above — The buffer, differential amplifier and low-pass filter employed during this series of experiments. Function is identical to the similar stage described earlier.



Above — The output of amplified and filtered Wheatstone bridge at open circuit (no resistance at the Unknown Impedance port).

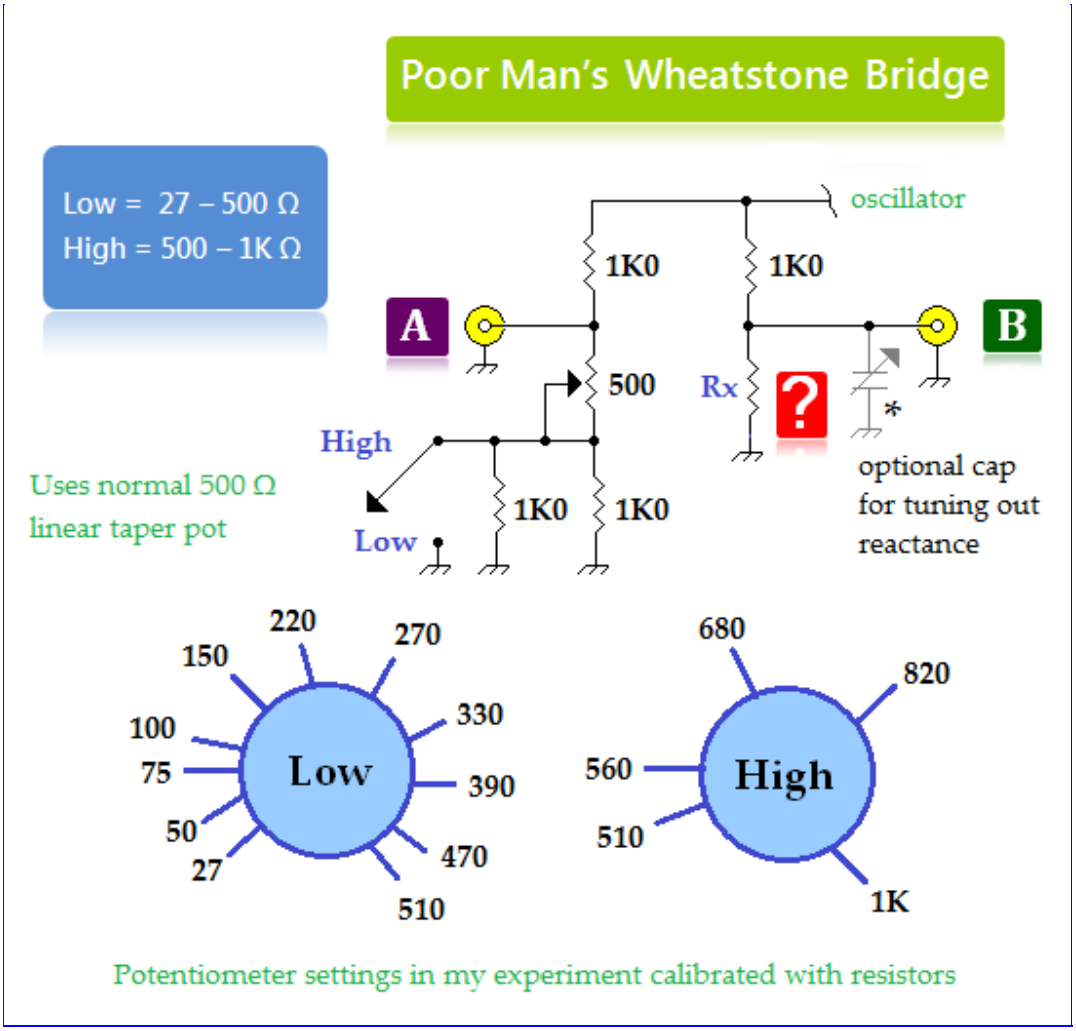


Above — the output of the amplified and filtered Wheatstone bridge at NULL (potentiometer setting balanced to match the resistance at the Unknown Impedance port. Below 2 mV, accuracy is lost.



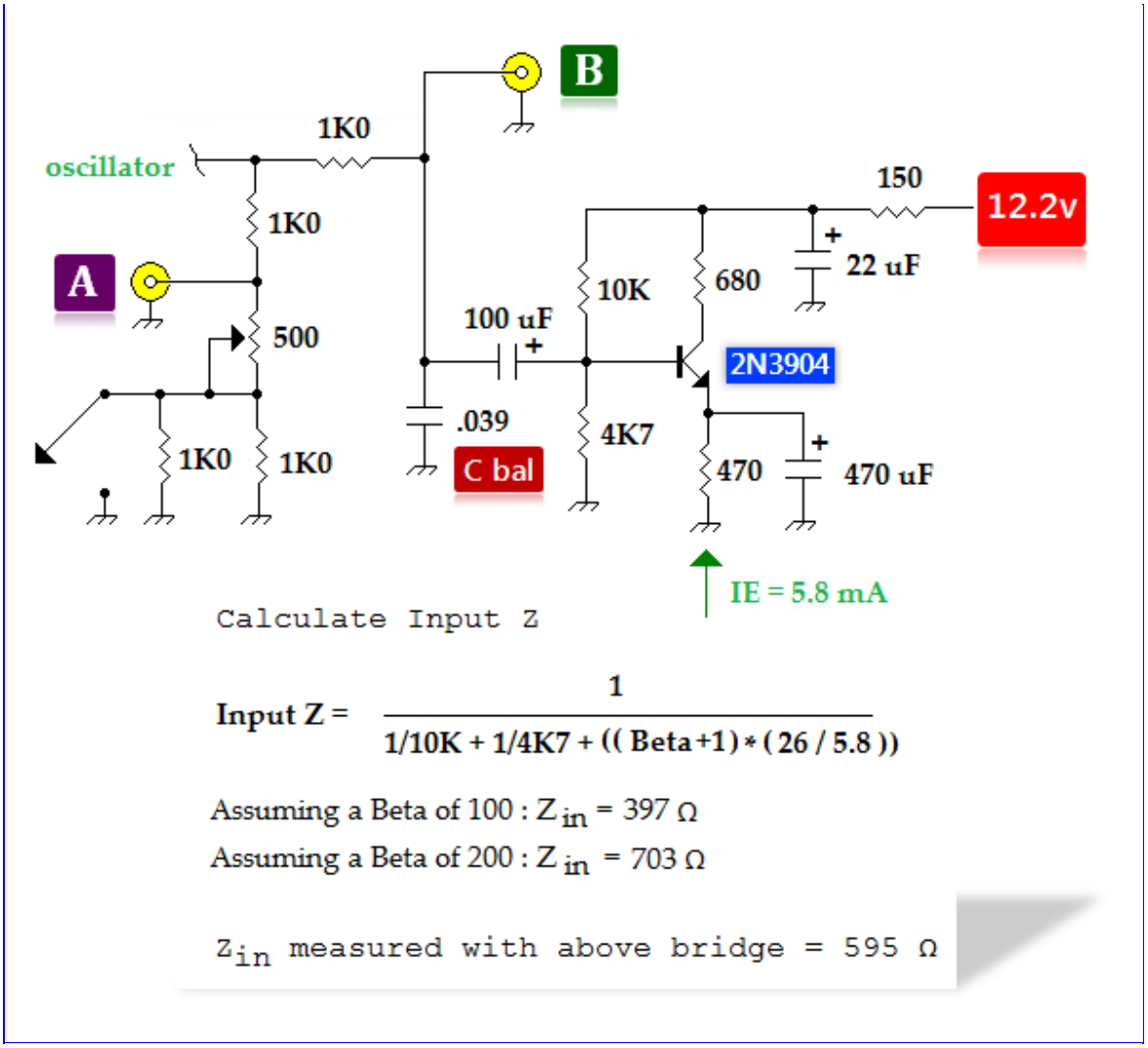
Above — The math behind the bridge. I found when R Variable (Rv) was the same resistance as the fixed resistor R parallel (Rp), reasonable resolution was possible. "Reasonable resolution" means your pot has a good range of rotation as you go from the lowest to highest measureable impedance. Generally, Rv has to be less than the maximal impedance you are trying to measure.

R Scale (Rs) can be switched in decades via a panel mount switch to cover a wide range of resistance with good resolution, or just be 1 or 2 values. It's your design call.



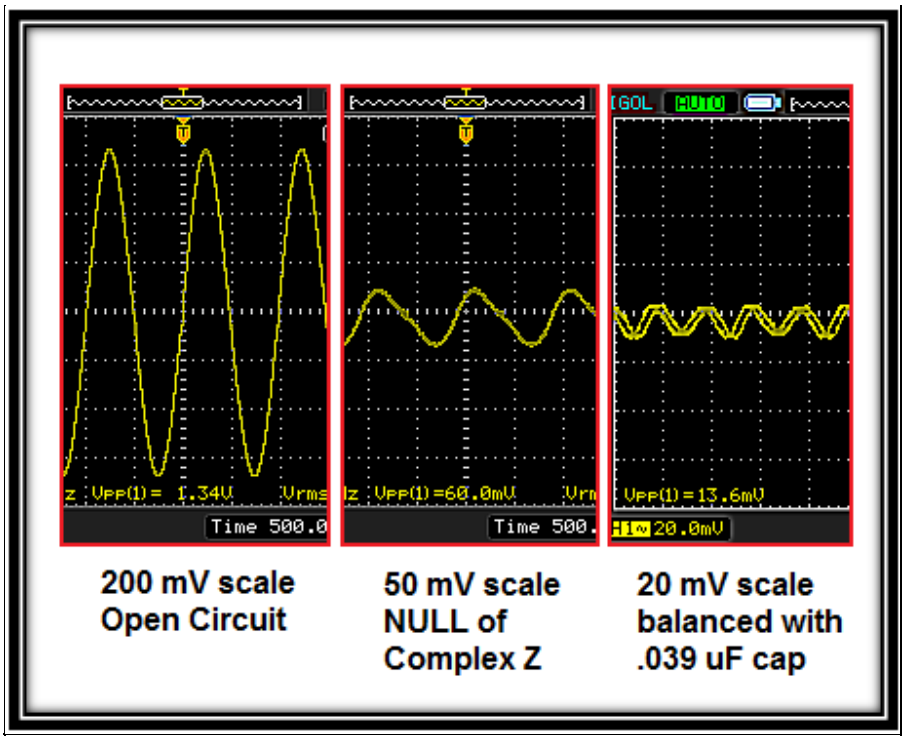
Above — My poor man's impedance measurement device that uses a common 500 ohm linear taper pot as the balancing resistor. In order to get good pot resolution, the desired range is switched. This bridge measures impedances at the Unknown Port from about 27 ohms up to 1K with decent resolution. The blue circles depict how I calibrated the front panel of my device using 2 colors. This device had an average return loss of 32.5 dB when a NULL was obtained.

Measuring resistors to calibrate a bridge is quite different from real-world measurement of reactive AF amplifier loads — if the unknown resistance has a large inductive or capacitive reactance, obtaining good bridge balance might prove difficult. Your bridge can *only* null the in-phase signal. An extension to the standard bridge involves adding a series or shunt capacitance (depending on the phase of the reactance) to the A or B port. This may allow you to null the reactive part and also provide the reactive impedance value as well. An outstanding reference may be located with your favorite search engine: Look for the manual for the **General Radio GR1650 Impedance Bridge.** I found a copy and the download was very slow, but worth it. This manual may be the greatest reference every published on the Wheatstone bridge and comprehensively covers tuning out the reactance of complex impedances amid a myriad of other topics.



Above — The poor man's bridge measurement of a test AF amplifier on my bench. The reactive component of the amplifier input impedance was minimized using a 0.039 uF capacitor found experimentally. Of particular interest, is the difference between the *calculated* input Z and the *actual* input Z measured with the bridge.

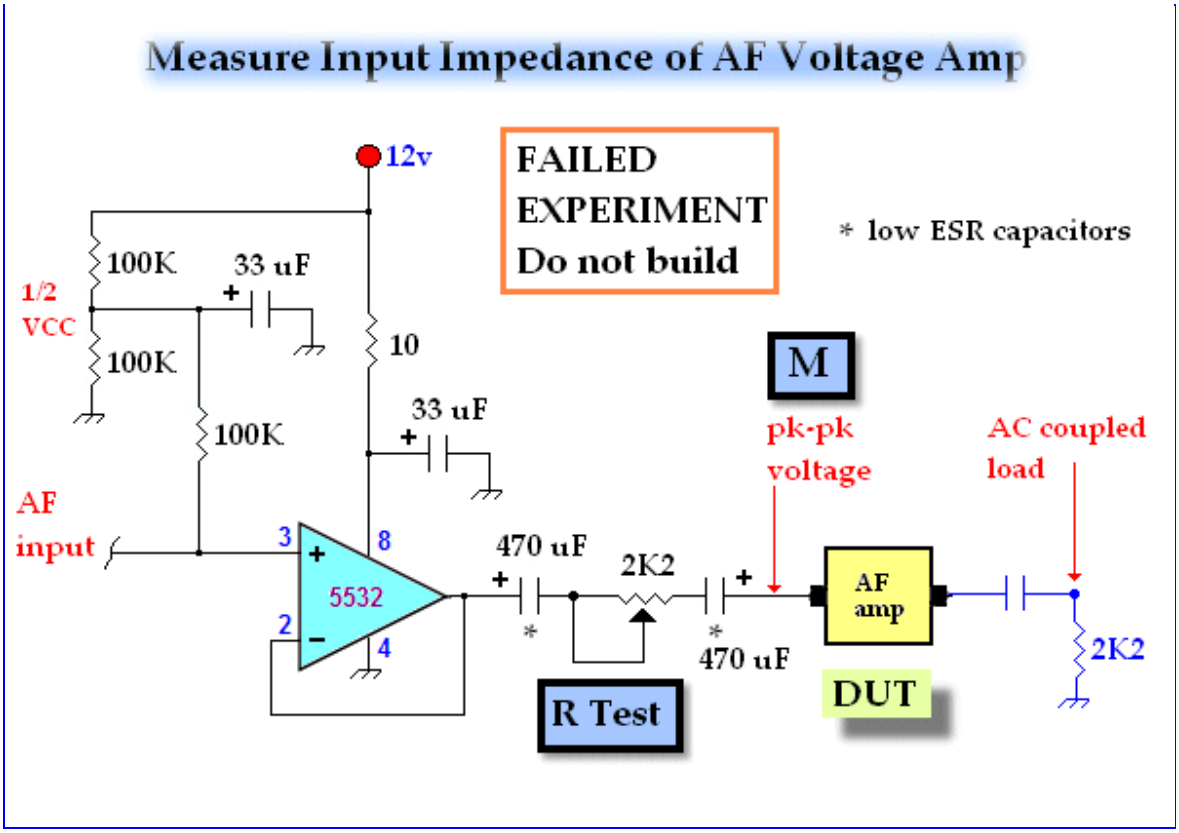
The Beta of the 2N3904's in my collection ranges from about 100 to 225. Calculations with 2 different Beta values are shown (RE is *well* bypassed with a 470 uF capacitor, so, re = (Beta * 26/5.8 mA) . The measured input Z was 595 ohms. I confirmed this by removing the 0.039uF tuning capacitor, plus connecting a fixed 595 ohm resistance to the Unknown Impedance port. I then turned the potentiometer fully clockwise and adjusted it for a NULL. When the bridge was nulled, the potentiometer knob pointed at the same mark as when the amplifier was connected to this Unknown Impedance port.



Above — The oscilloscope output waveform of the amplifier circuit shown above: open circuit, with the potentiometer balanced as well as possible and finally, the potentiometer balanced with the addition of a 0.039 uF capacitor attached to Port B. Although, I was able to get a NULL without the capactitor by just tweaking the potentiometer, slightly better precision was obtained after adding the 0.039 uF balancing capacitor.

Part 4: Miscellaneous Circuits, Scans and Photographs

Above — More accurate results will be obtained with a calibrated 10 turn potentiometer to balance your bridge. A local store sold me this precision 10-turn, 10K pot for about 11 dollars (still expensive for me). They normally sell for twice this price in Canada.



Above — My first input impedance measurement device that didn't work. It turns out my experiments were performed *incorrectly*, however, I'm glad because this failure spurred me to investigate bridge networks. The series resistance method *is* worth understanding and happily, Jeff, AD6MX described the correct procedure in a private email received December 2010. I quote him below:

"The series resistance method for input impedance should start with the variable resistor disconnected from the node to be measured. The open circuit voltage at the end of that resistor is measured (the resistor value doesn't affect the open circuit voltage since there's no current into an open circuit.)

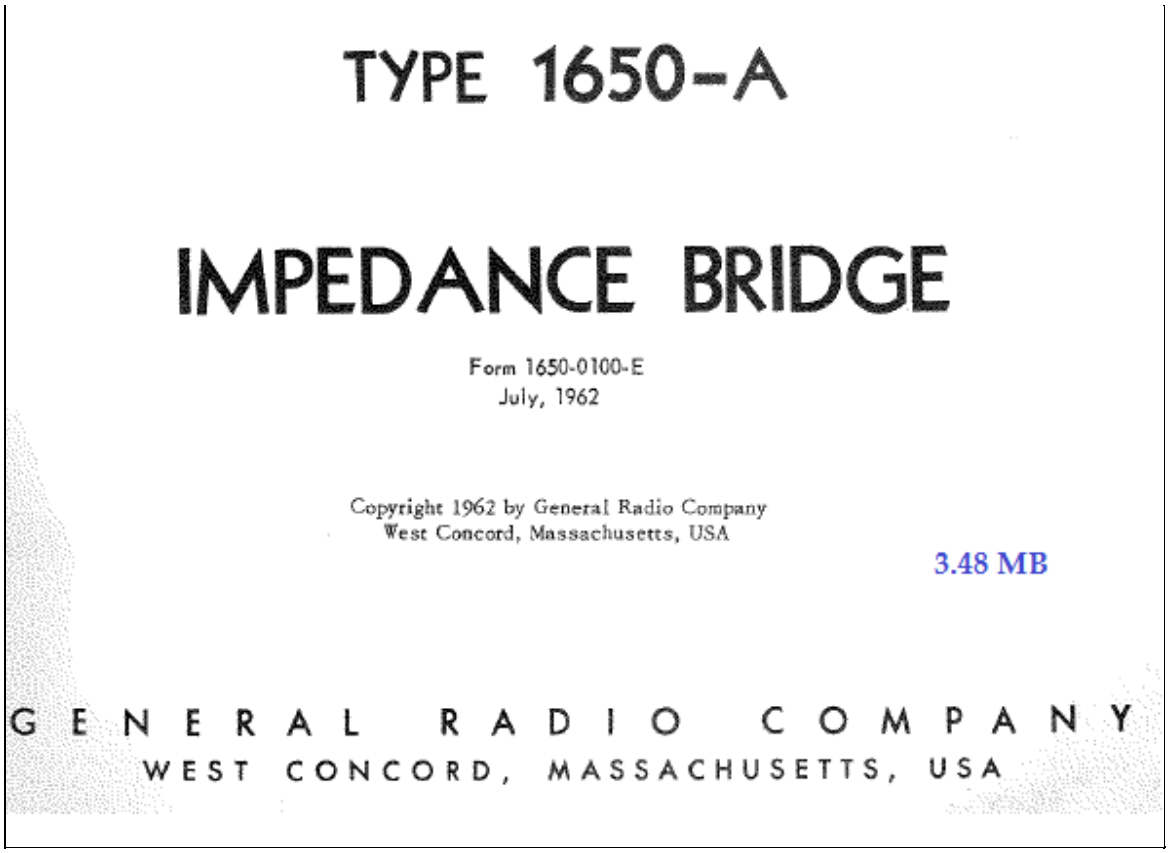
Next the free end of the variable resistor is then connected to the input node and the resistor is finally adjusted for half the open circuit voltage at the same end of the resistor, at the input node being measured. What happens is the variable resistor and the node input impedance form a voltage divider, with equal arms or branches.

The value of the resistor when measured out of the circuit is the same as the input impedance at the measured input node. This scheme has some assumptions: the driving amplifier has negligible output impedance compared to the measured impedance, and the input impedance is purely resistive, with no reactance or V-I phase shift.

The phase shift condition may be checked by taking these 3 voltage measurements: across each branch of the divider separately, and also the driving source voltage (across both branches in series.) The sum of the separate branch voltages should match the source voltage when there is little phase shift.

This 3 voltage scheme is used in some antenna analyzers in order to measure phase shift. For checking for the resistive condition, it's not important the 3 voltage method has a sign ambiguity which needs an additional step to resolve. Your description seemed to suggest starting with zero series resistance, but you see that is not the same as the procedure above. The applied voltage needs to be small enough that the amplifier remains operating in its linear range during the measurement".

Thanks for this info Jeff!



Above — A modified scan of General Radio's *über awesome* manual for the Type 1650-A Impedance Bridge.

RF Workbench Page 3

This web page is the third installment of a 6 part series that explores basic measurement of RF circuits.

Part 3, further examines Return Loss Bridges from a bench-practice viewpoint.

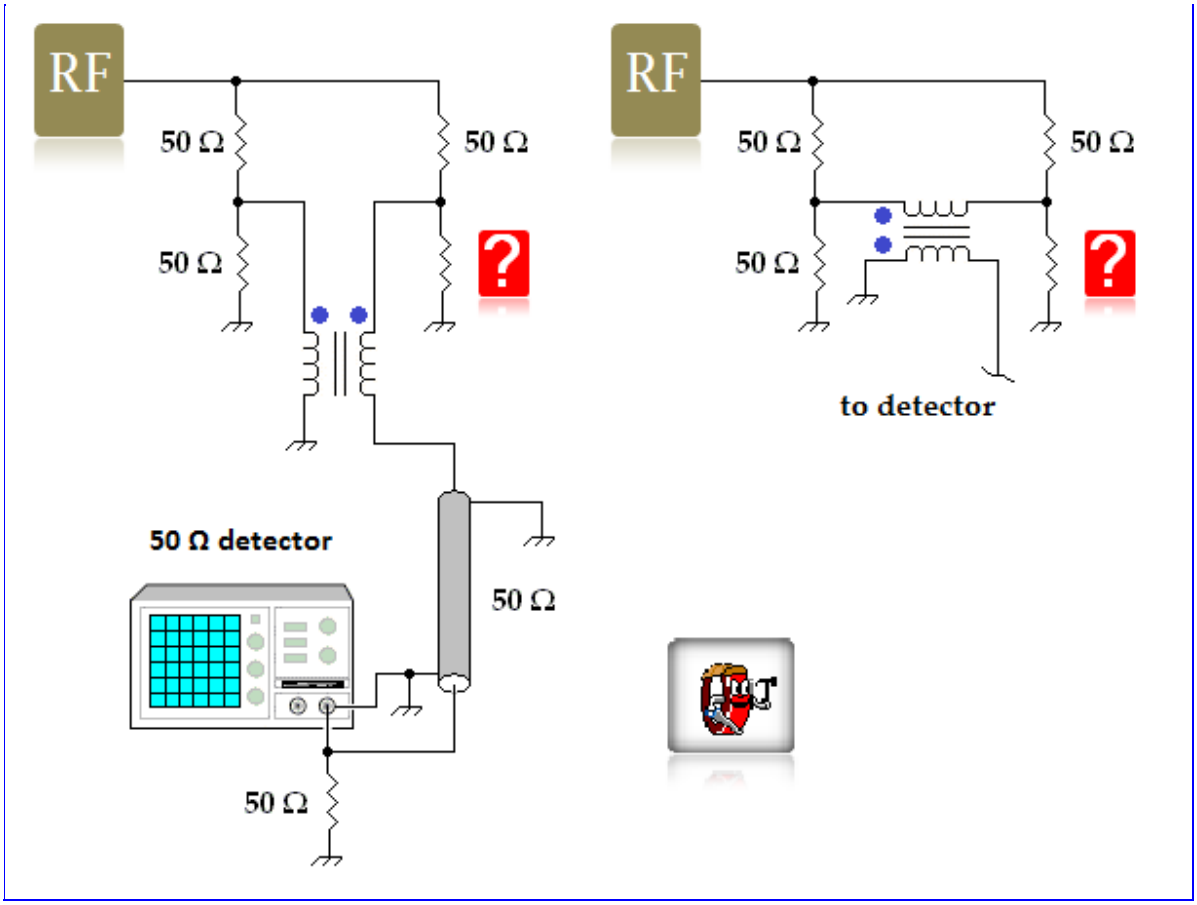
I borrow heavily from the work of Wes, W7ZOI per correspondence, direct contributions and from EMRFD.

I focus on measuring low-level, HF circuits with a return loss bridge — topics such as using the bridge for antenna matching are omitted and readily found on the web. This web page contains minimal text and just relies on simple diagrams and photographs to transfer ideas and knowledge. Information regarding wideband bridge network function may be found elsewhere on this and other web sites and in EMRFD.

More on Return Loss

Gear

Equipping for a 50 Ω measurement environment in 2010 greatly improved my design capacity. The 50 Ω terminated oscilloscope makes a sensitive and accurate detector for return loss measurement. Discussion about using a 50 Ω oscilloscope termination is on the RF Workbench 1 web page.



Above left — The RLB and measurement set up from EMRFD. Occasionally, you may see bridges using a different balun transformer wiring as shown to the right of this figure.

Above — All the needed parts to home build a return loss bridge. For some, the parts investment might seem substantial, but what hobby isn't expensive? If you consider the cost of commercially manufactured bridges, a homebrew solution seems a bargain. Recycled parts and a home-built chassis are inviting cost-containment techniques. See the web site of Jim, K8IQY for an example of a homebrew RLB chassis. Jim, a Manhattan style construction wizard, builds the nicest looking gear — he puts me to shame.

Above — A completed bridge. I used 1% tolerance 49.1 ohm resistors and an FT50-43 ferrite toroid for the bifilar wound transformer. Inductance = 38.4 uH. Many builders use the FT37-43 ferrite core. I prefer using 2 colors of enamel coated wire to avoid confusion when building stuff with transmission line style transformers and all I had in 2 colors was 24 gauge wire The bigger size ferrite toroid better accommodates the 24 gauge wire, plus photographs better.

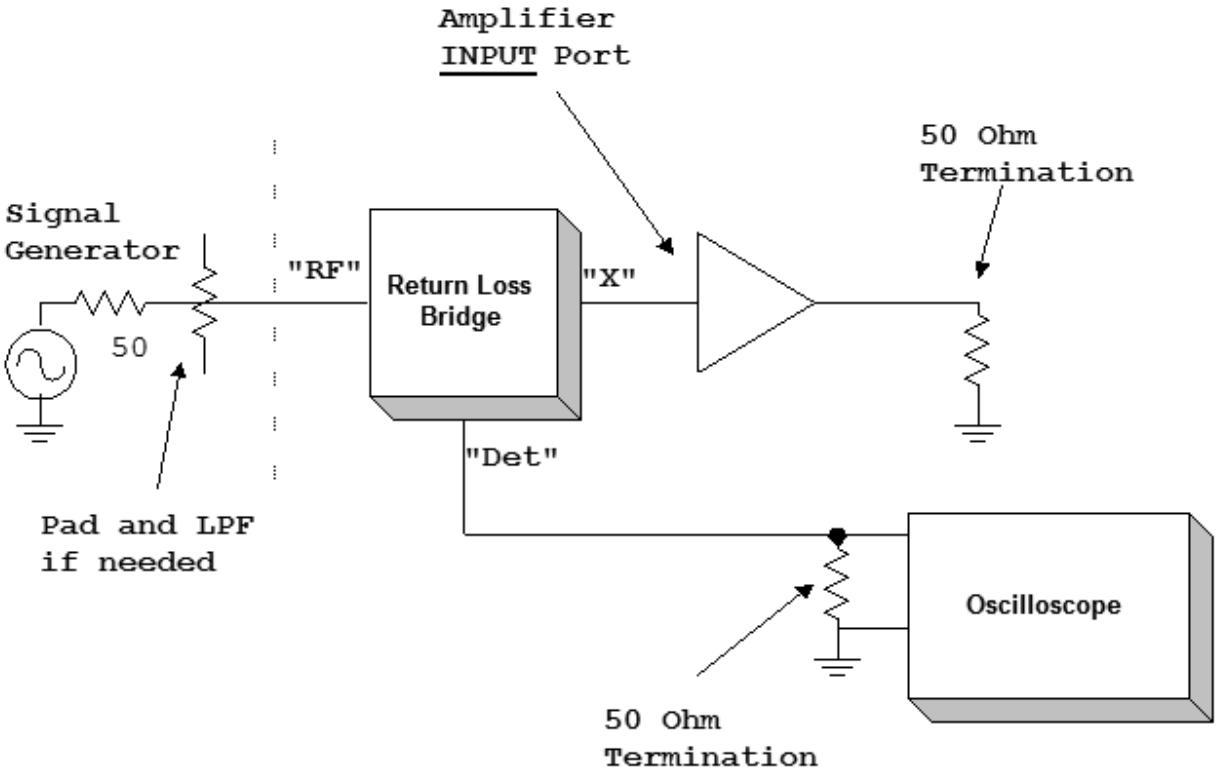
Bridge directivity of the above RLB was 30 dB at 7 MHz, 34 dB at 14 MHz, 35.6 dB at 21 MHz, 42.1 dB at 50 MHz and 43.4 dB at 100 MHz.

If you build a circuit with a return loss close to 30 dB, it's a good day.

Above — The completed RF-tight bridge. Don't forget to label your network ports.

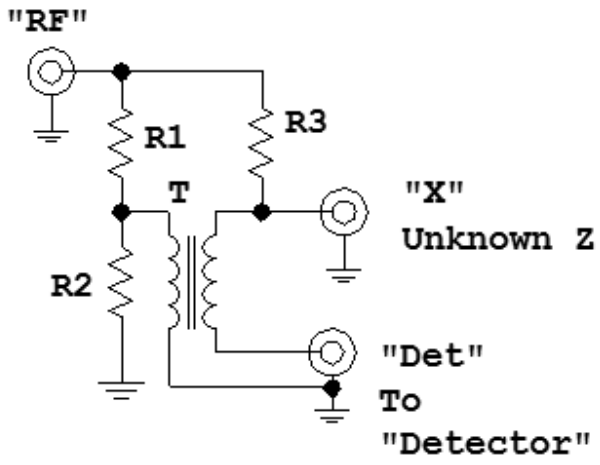
Measurement Technique

Measurement of an Amplifier Input Return Loss



Measurement Procedure:

1. Terminate X port in an open circuit and note level. e.g.,250 mV pp at scope.
2. Attached a 50 Ohm load at X just to check the bridge. e.g., now see 5 mV pp.
3. Bridge directivity is 20Log(250/5)=34 dB. This is the best you can measure. Better numbers are meaningless.
4. Remove the 50 Ohm load. Attached the amplifier input and note level. e.g. 18 mV pp.
5. Amp Input RL is 20Log(250/18) =23 dB.



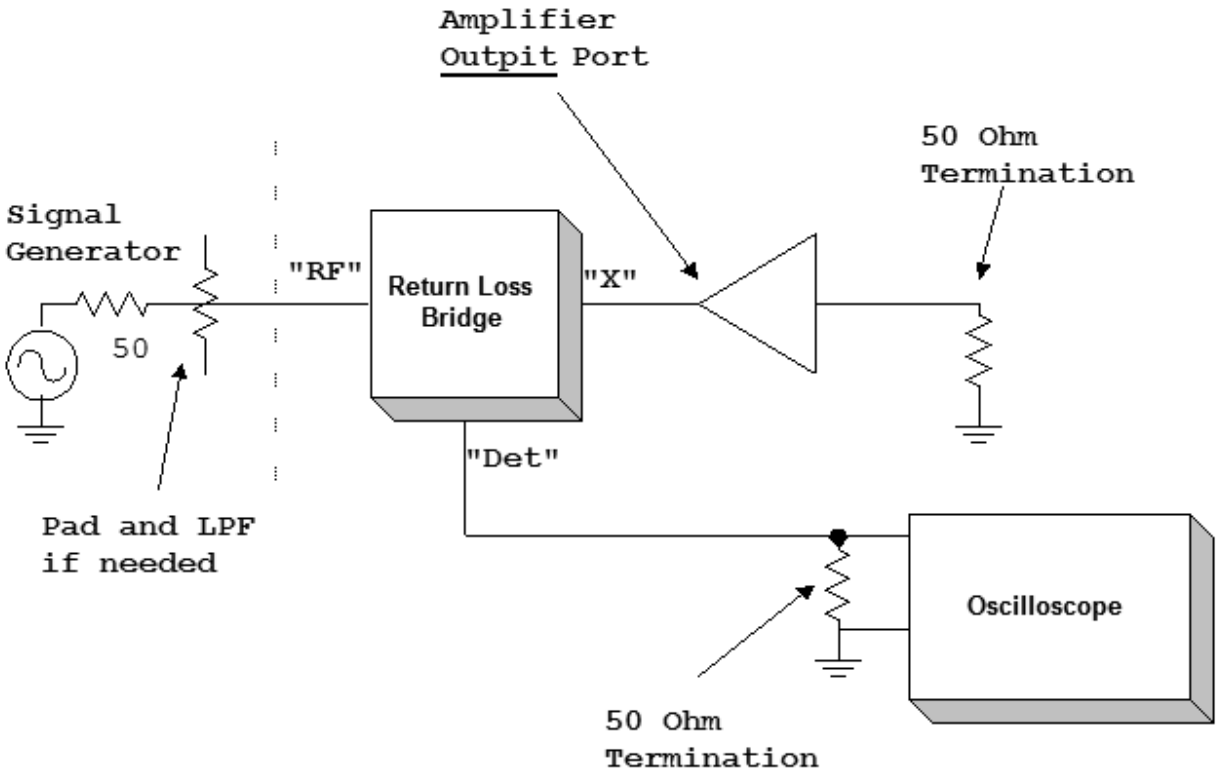
24Dec10
w7zoi

Above — Measure the return loss input of an amplifier. You'll need at least one 50 ohm BNC feed-through terminator on your bench to test amplifiers with wired-in BNC connectors (such as on this amp); else just solder a 47, 49.9 or 51 ohm resistor from the amplifier output to ground. The BNC connectors allow you to quickly and solderlessly interface components such as filters, attenuators, oscilloscopes or **50 Ω** signal generators.

A typical amp measurement work flow may go something like this: Measure gain using a signal generator and the 50 ohm terminated scope; add the bridge and measure input return loss; finally, flip the amplifier around and measure output return loss. All 3 functions can be performed in 2

— 5 minutes including time to drink coffee + perform calculations by computer or with a HP scientific calculator.

Measurement of an Amplifier Output Return Loss



Measurement Procedure:

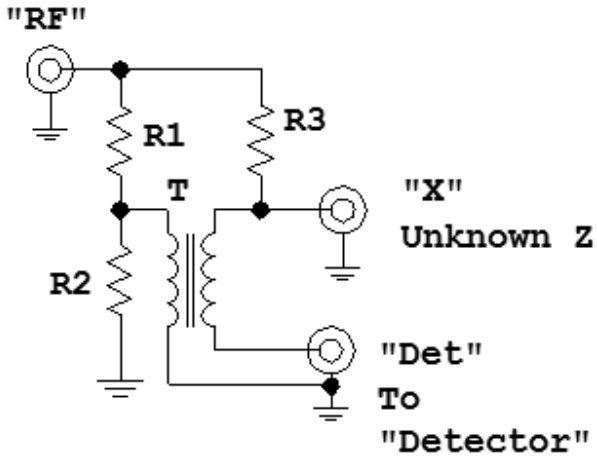
1. Terminate X port in an open circuit and note level. e.g.,250 mV pp at scope.

2. Attached a 50 Ohm load at X just to check the bridge. e.g., now see 5 mV pp.

3. Bridge *directivity* is 20Log(250/5)=34 dB. This is the best you can measure. Better numbers are meaningless.

4. Remove the 50 Ohm load. Attached the amplifier output and note level. e.g. 32 mV pp.

5. Amp Output RL is 20Log(250/32) =18 dB.



24Dec10
w7zoi

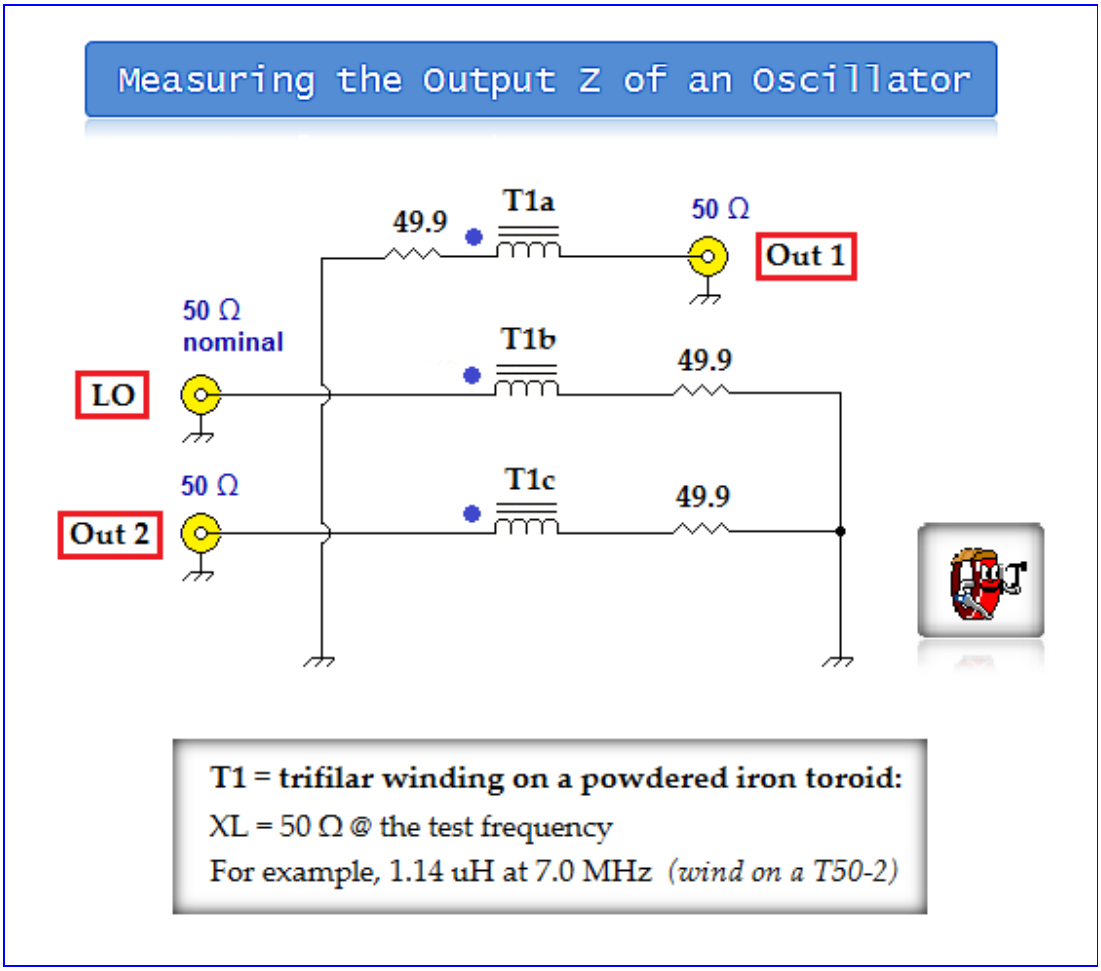
Above — Measure the return loss output of an amplifier. The above 2 procedural diagrams provided by Wes, W7ZOI. Many thanks to Wes. These figures are copyrighted © by Wes Hayward, 2010.

Your signal generator should have a return loss of at least 20 dB for greatest accuracy — all of my bench test generators have at least 30 dB of return loss. If you have a signal generator with a low impedance output and place a 10 dB attenuation pad on the output, you'll have at least 20 dB of return loss.

In the above figures, Wes gives an open circuit return loss of 250 mV; I set my signal generator output so the open return loss is somewhere between 170 and 250 mV; this allows you to accurately measure a really good 50 ohm return loss at >= 5 mV or so. Some people may have trouble going any lower than 5 mV due to scope accuracy. This is just something to consider.

Bench Exploration

For me at least, a special case of return loss measurement exists; measuring the return loss of a local oscillator. Since the oscillator under test must be *on* during measurement, it's emitting a signal at the same frequency as the bridge signal generator and interferes with measurement. If some 50 ohm attenuation is added to reduce the local oscillator under test output signal amplitude, this increases the return loss of the local oscillator under test. This is normally a good thing, however, we seek the **raw** output return loss or output impedance of the local oscillator under test.



Above — An initial experiment that a builder from Michigan, USA and I first used to measure the output impedance of a local oscillator consistent with the breadth and scope of this web site. We wanted something simple and wished to avoid building a vector network analyzer or performing ugly algebra. I built a simple crystal oscillator for 7.0 MHz using an output transformer wound to give a low impedance output. The circuit was measured and calculated using the instrument above and the formula and procedure below.

The calculated output Z was 33.2 ohms. I build a standard value resistor 6 dB attenuator pad from this table. After fitting the pad, I re-measured and re-calculated the output impedance at 46.8 ohms. This seemed okay. I built a couple of oscillators for other frequencies and the output impedances were hundreds of ohms! — disappointing. Still, we were on the bench in a solution-focused mode and needed to try something else.

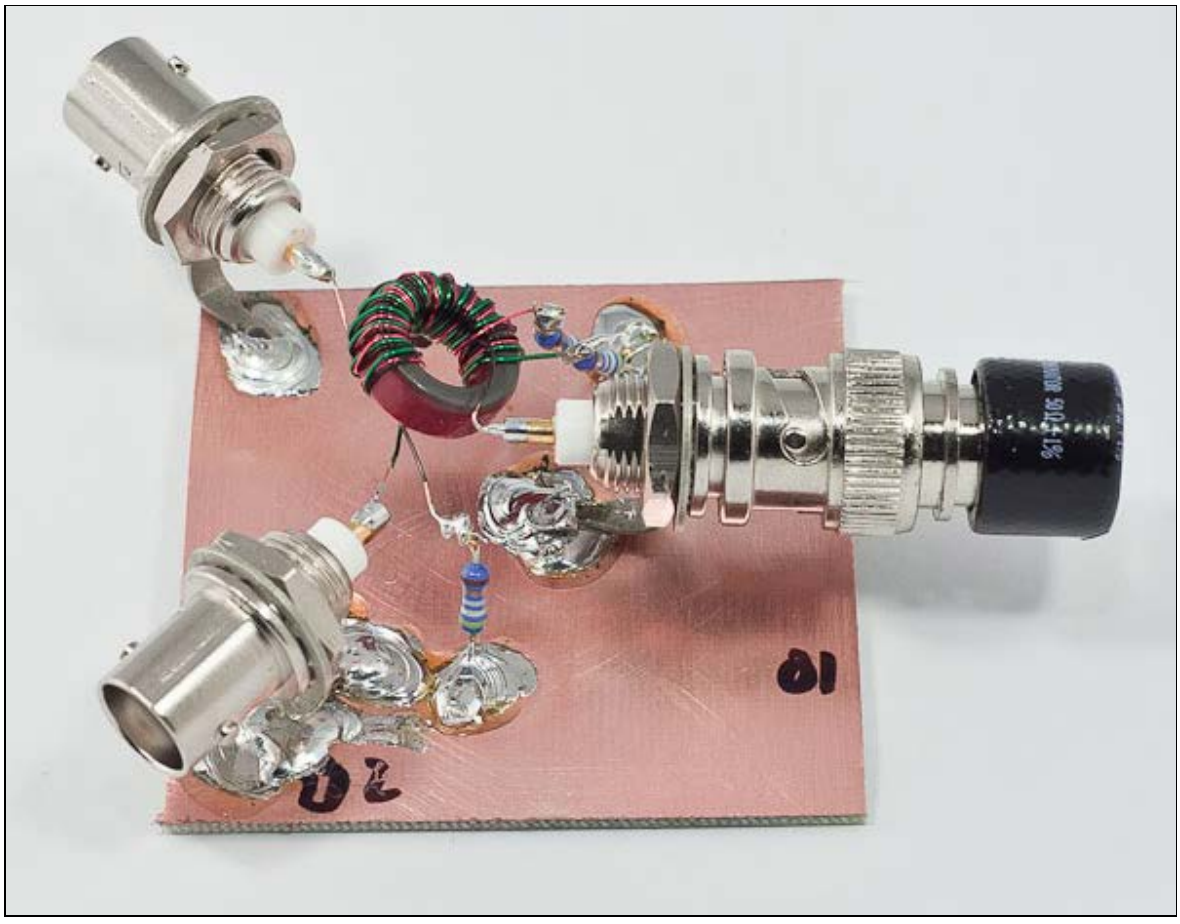
Formula

$$\text{LO output Z} = \left(\frac{\text{Out1} + \text{Out2}}{\text{Out1} - \text{Out2}} \right) \times 50$$

Make Out1 the larger pk-pk voltage

Out1 and Out2 are connected to 50 ohm terminated scope inputs, or 1 of them is connected to a 50 ohm terminated scope output, while the other is terminated with a 50 ohm resistor.

Above — The formula for calculating the output impedance with the *experimental* local oscillator output Z device.



Above — Breadboard of the experimental L.O. output impedance bridge with a 50 ohm feed-through terminator on the Output 1 port. It failed to work as expected. Skillful adult problem solving goes something like this:

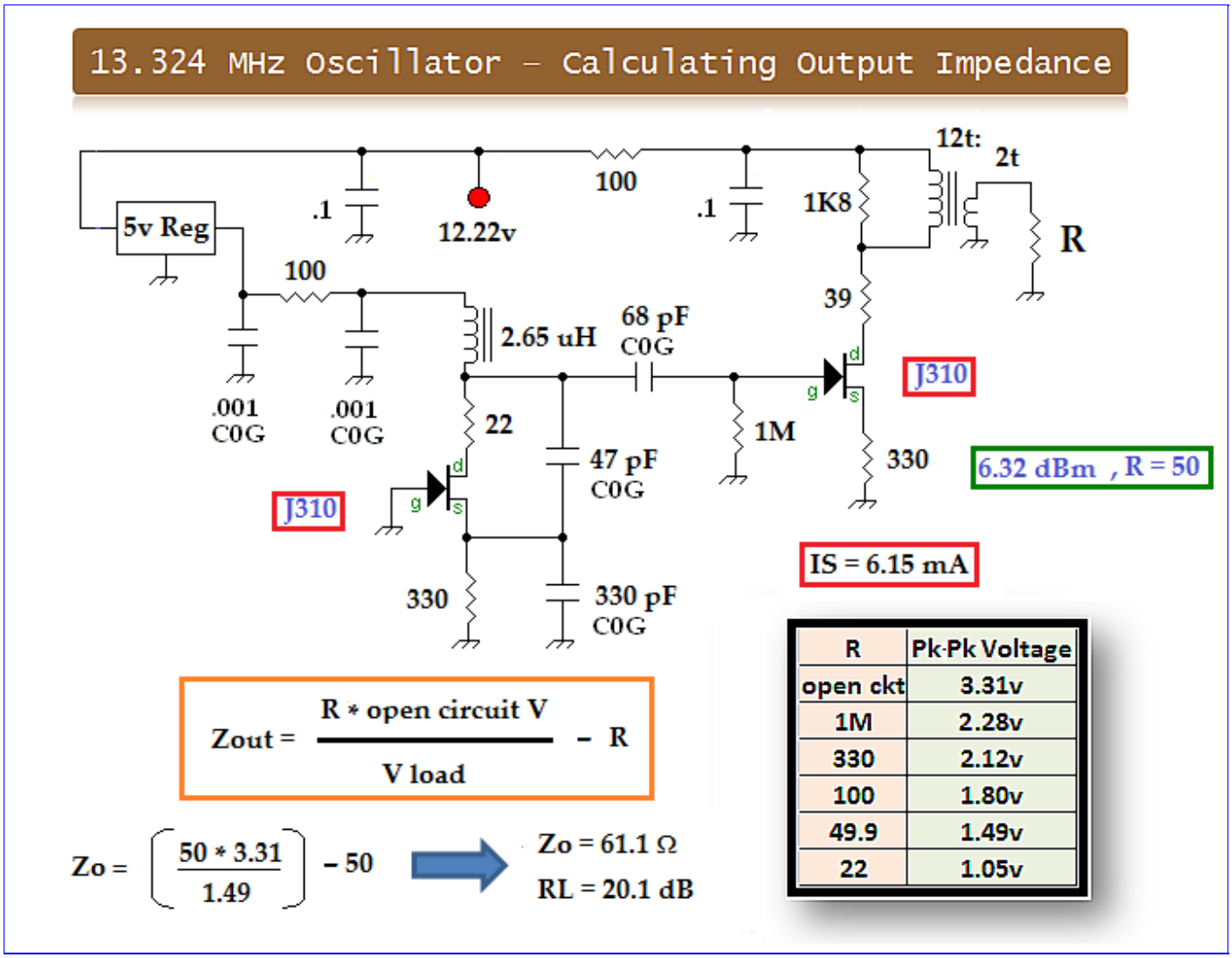
1. Identify the problem.
2. Brainstorm to generate some potential solutions.
3. Try out one of your ideas.
4. If that doesn't work, try another idea.
5. If none of your ideas work, wait a while, or ask an expert.

Well, we ran out of ideas and decided to ask experts for some more ideas; Professor Kuhn and Wes, W7ZOI.

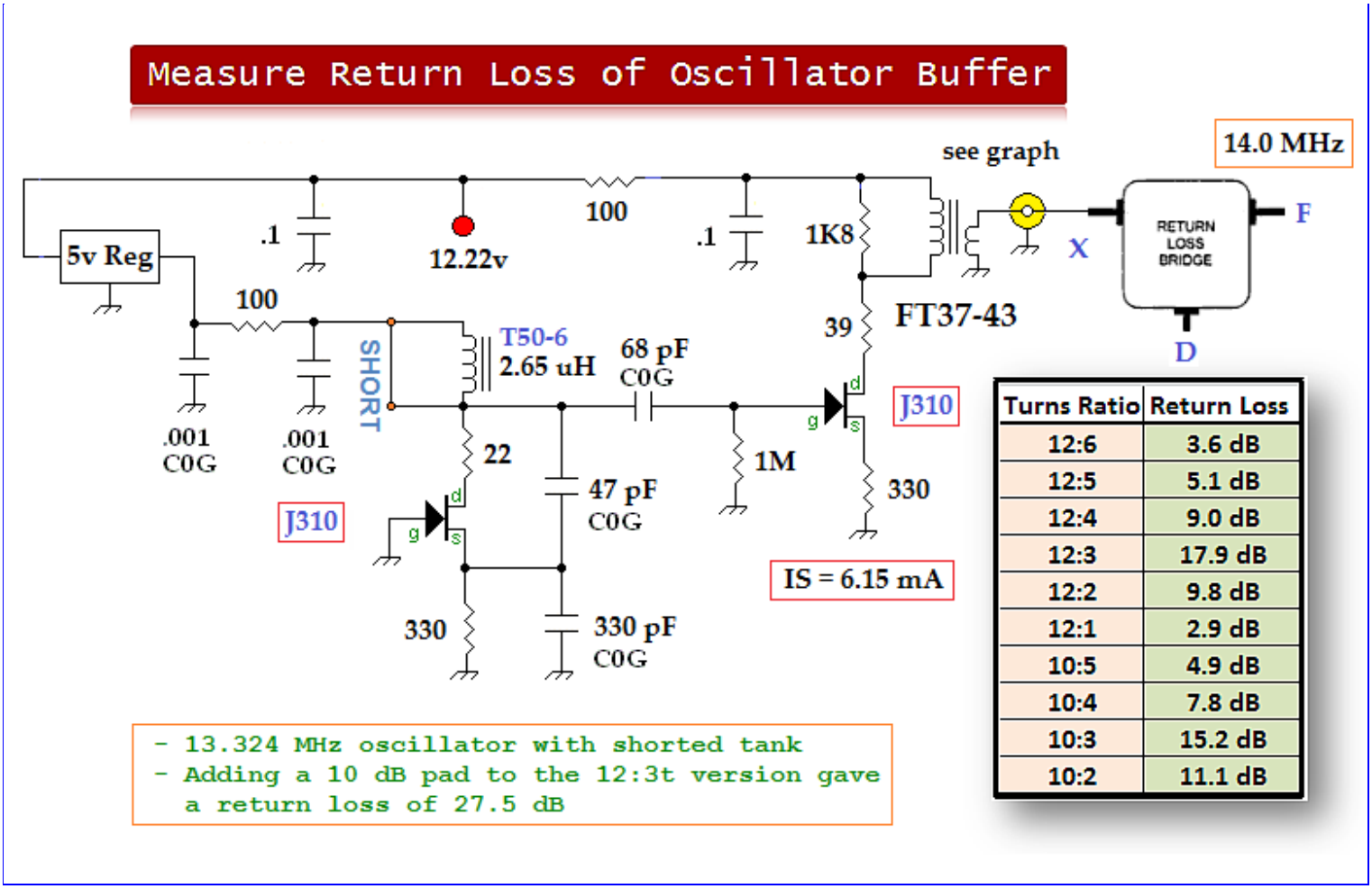
I'll share their key messages. First, accurately measuring the output impedance of an RF oscillator can be difficult — measuring the return loss of a buffer amplifier is much easier. For this, some builders run the bridge signal generator on a slightly different frequency than the oscillator under test while using a spectrum analyzer as the 50 ohm RLB detector.

Another way is to short circuit the tank on the oscillator and measure the buffer output in the normal way — a popcorn solution indeed!

We tried calculating oscillator output impedance using different equations and 1 example is shown below. Failing to account for inductive and/or capacitive reactance plus resistance in the output circuit (including the transformer), plus upsetting the circuit during AC voltage measurement adds uncertainty to calculations — measurement seems more reliable.



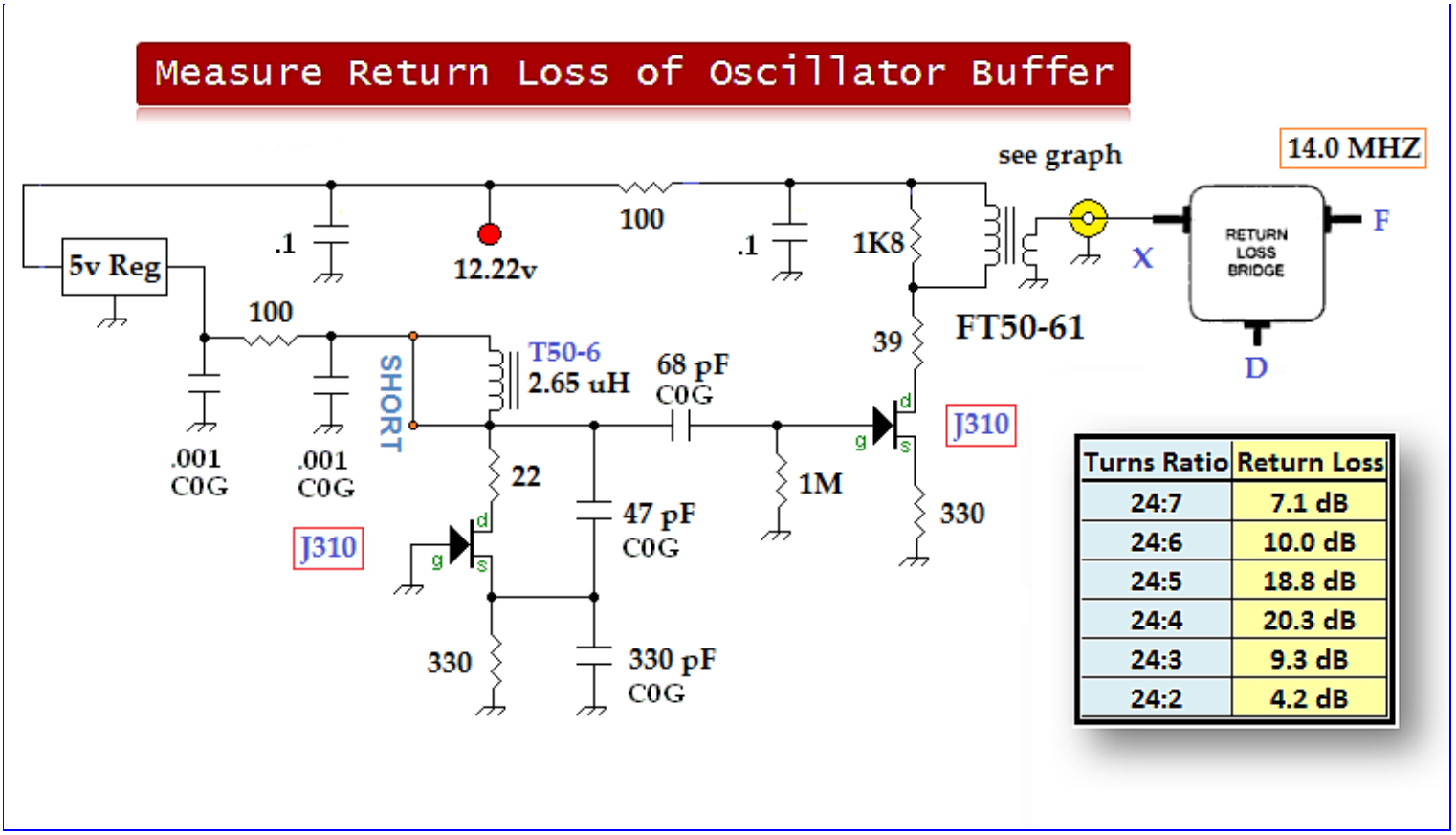
Above — One method of calculating output impedance. Running the output at open-circuit likely effects the oscillator by changing its load despite having the JFET buffer. Some builders use this equation for calculating the output impedance in their audio amplifiers. This amplifier **should** have a 50 ohm output impedance based upon the transformer turns ratio and the 1K8 resistor across transformer: 1800:50 ohms = a 36:1 impedance ratio and a 6:1 turns ratio.



Turns Ratio	Return Loss
12:6	3.6 dB
12:5	5.1 dB
12:4	9.0 dB
12:3	17.9 dB
12:2	9.8 dB
12:1	2.9 dB
10:5	4.9 dB
10:4	7.8 dB
10:3	15.2 dB
10:2	11.1 dB

Above — A simple 13.3 MHz L-C oscillator was built and evaluated. After shorting the tank coil, return loss versus turns ratio was measured and tabled as shown. To my surprise, I observed the best match with a 4:1 turns ratio. This suggested that the transformer, wound on a ferrite FT37-43 toroid was exhibiting high resistance and far from the "ideal transformer". The inductance of my 12 turn transformer was 38.3 uH.

The initial secondary winding had 6 turns and then was reduced sequentially by 1 turn. After removing each turn, the 1/2 cm of increased wire length was cut off and the enamel scraped off of the new wire ending to ensure a short connection to the output jack. During measurement, unless the secondary transformer wires were kept tidy, a ~40 MHz oscillation occurred when the 1K8 resistor was disconnected. The 1K8 resistor prevented such oscillations and improved the return loss by 1-4 dB at the various turns ratios. Testing frequency was 14 MHz.



Above — The same circuit with a lower-loss, FT50-61 ferrite transformer. I could have used a FT37-61, but prefer the 50-61 as the bigger core allows the use of heavier wire which provides some robustness when performing intensive experiments. The inductance of 24 turns on a FT50-61 measured 36.2 uH. Although lower permeability ferrite toroids require more windings, this transformer is closer to the "ideal transformer" than that wound on a FT37-43 ferrite core — a 6:1 turns ratio gave the best return loss; the output Z is pretty close to 50 ohms.

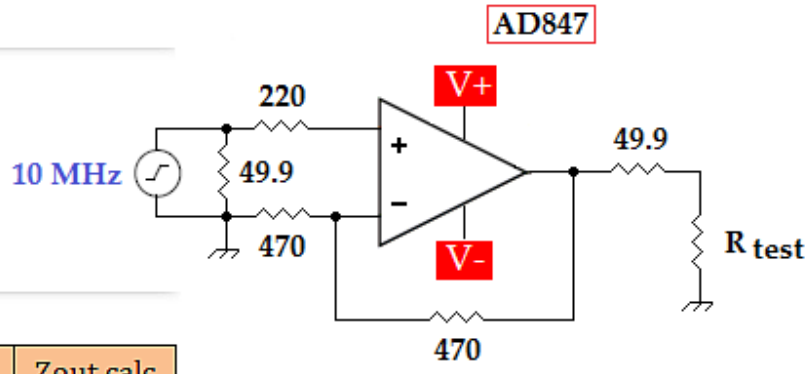
The information garnered during these tests proved enlightening and reinforces why bench measurement provides the greatest way to learn about and optimize your circuits. I hope this simple web page on return loss measurement fuels your own experiments — the most important experiments will be those *you* do on *your* bench.

Miscellaneous Figures and Photos

Calculate Output Z of an Oscillator – Experiment 1

Measure the open circuit output amplitude of the oscillator and then add a load so that the loaded voltage dropped to something in the range of 50 to 75 percent of the open circuit voltage. Then use voltage divider theory to calculate the source resistance of the oscillator.

$$V_{loaded} = \frac{V_{unloaded} * R_{load}}{(R_{oscillator} + R_{load}}$$

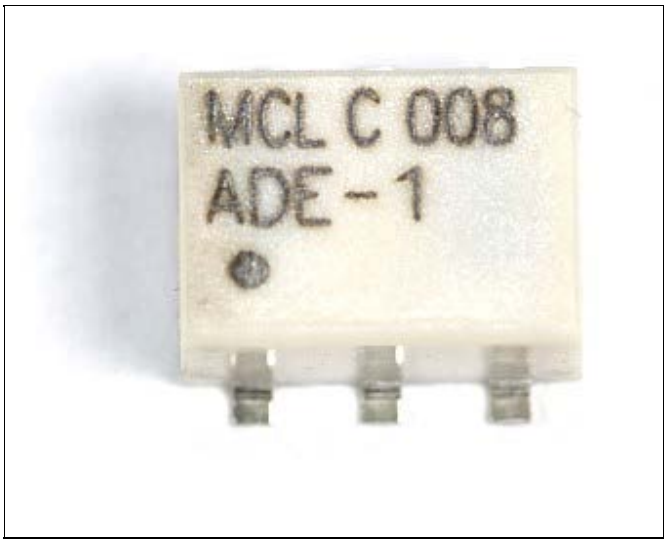


Rtest	Vout	Icalc	Pcalc	Zout calc
1000000	0.9999	9.999E-07	9.998E-07	
220	0.8148	0.0037	0.0030	49.9779
50	0.5003	0.0100	0.0050	49.9300
22	0.3056	0.0139	0.0042	49.9823

RF — Test and Measurement

Miscellaneous RF Topics 2011

Introduction



My basic goal for Fall and Winter 2010 was to fearlessly advance my RF design ability — pushing just beyond my comfort zone to impose the psychological stress that promotes focused learning. You may have experienced this in University or when working against a deadline. Cramming, "burning the midnight oil", or locked room brainstorming exemplify this approach.

These circuits feature carefully measured input and output impedances (nominal impedance = 50 Ω), plus voltage gain and DC current.

This web site spawns email with builders worldwide. Our interactions are varied; getting help, giving help — or just chat. On occasion, I design, build and/or test circuits to help struggling builders. After spending considerable time, I email them my work hoping it will help. Often enough, I never receive any acknowledgement from these readers — did the circuit work or did they appreciate I spent 1-2 hours researching their concern?. This is actually normal — we must constantly strive to overcome our innate, self-centered nature; lest it dominate our behavior. все нормальные.

To that point, I wish to gratefully acknowledge the people who support me in this hobby: Wes, Ken, Scott, Peter, Tom and the many others whose email advice and published and private work informs and inspires me.

Topics:

1. Transmit Mixer Experiments
2. Bipolar Transistor Feedback Amplifier Experiments
3. JFET Common Gate Transistor Amplifier Experiments

Navigation and Preamble:

This web page grew into a large monster — and includes a supplemental web page with numbered topics referenced in the text. I apologize for the navigation difficulties this web page poses. Equal time was spent experimenting with the circuit designs and circuit photography. I strive to provide a variety of bitmap and photographic image styles on this web site.

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1. Transmit Mixer Experiments

Since I've never experimented with transmit mixers, I didn't appreciate how much time goes into their design. Consider, for example, the LO system from the project entitled A Monoband SSB/CW Transceiver in Chapter 6 of EMRFD. The mega low (about -20 dBm) output from a diode ring mixing a VFO and crystal oscillator is triple tuned band-pass filtered and then amplified to +8 dBm. Continuing on, the transmitter chain features more mixing, band-pass filtering and voltage amplification by a feedback amplifier chain boosting the signal to around 300 mW. The circuits needed to mix, filter and amplify this RF chain would challenge most amateur designers — me included.

Contrast this with a typical first transmitter built by a new builder. Likely your first scratch homebrew transmitter consisted of a crystal oscillator, a keyed Class A buffer/amplifier and perhaps a Class C final amplifier. No mixer was needed for we obtained a crystal cut on the frequency of

choice. Our focus was **power**— getting 0.25 to 1 watt into our antenna system! A good example was the Tuna Tin 2 transmitter by the late Doug DeMaw, W1FB that only used 2 stages. Although Doug wrote his 1976 article for Hams to build a transmitter from parts found at home, kitted versions are sold today.

Returning to transmit mixers — as amateur designers, we likely need to start on a small portion of the transmit chain and then after developing some competency, slowly extend our experiments all the way to the antenna port. In Fall 2010, I just examined some basic transmit mixing to get a feel for what's involved and what to expect. Mixing signals is a complex affair encompassing topics such as intercept point, conversion gain or loss, image noise suppression, noise figure, spurious/intermodulation products and port isolation. To keep things simple, only mixer port isolation and reducing spurious mixer products were examined.

Before beginning, I express the following concern: We experimenters, as stewards of the airwaves, must build exemplary transmitters with very low spurious outputs. I follow the example of Wes, W7ZOI and others — my transmit chains have spurious frequencies at least 50-60 dB down from the carrier (dBc). As a web author and radio amateur, I never want to directly or indirectly contribute to RFI and hope you agree.

Why Use a Transmit Mixer?

If you plan to design a superheterodyne based transceiver, you'll probably need to use a transmit mixer. Also mixing 2 frequencies permits using cheap microprocessors crystals to target a desired transmit frequency; separate crystal oscillators drive the RF and LO ports of the mixer. For added flexibility, the LO can be converted to a VFO once you have the basic design working well.

I purchased a bag of low cost crystals. By mixing 2 appropriate crystals, output on a Ham band is possible. For example, crystals at 2.048 + 5.0688 MHz = 7.117 MHz; 4.194 + 11.228 MHz = 7.034 MHz; and 3.932 + 11.046 MHz = 7.114 MHz. I frequently operate QRP on 40 Meters in the USA Novice band, so 7.114, or 7.117 MHz is okay. This helps CW operators avoid all the RTTY and QRM down in the traditional 40M band QRP frequency window.

Some Mixer Bullets

Mixers have 3 conventionally named ports; RF, LO (local oscillator) and IF (output).

The diode ring mixers presented are Level 7 mixers. Maximal LO power is 7 dBm.

Many builders limit the maximum RF power into a Level 7 diode ring transmit mixer RF port to between 0 to -3 dBm.

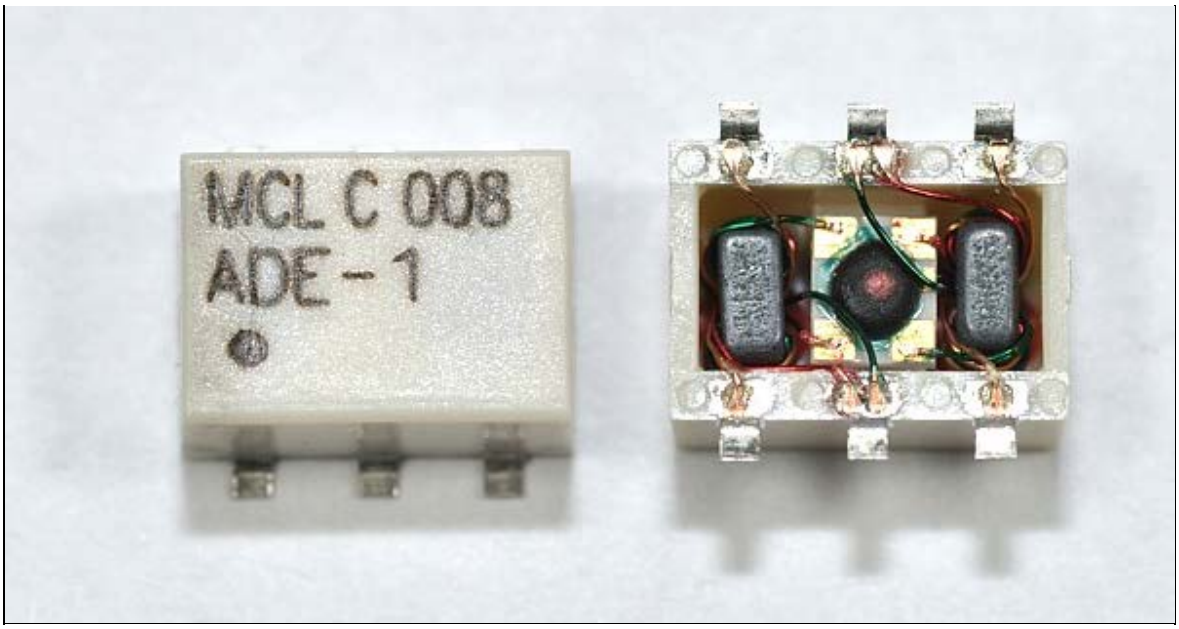
The term *isolation* refers to the amount of LO power that leaks into the RF or the IF ports.

Low-pass filtering the LO *can* significantly reduce harmonic products in a mixer

The top of the spectrum analyzer screen (always the top, and never the bottom) is called the reference level. That is the power at the top. If you have a signal generator with the output adjusted to be -27 dBm and pass this signal into the spectrum analyzer and adjust the attenuation in the analyzer to put that signal at the top of the screen, you then the reference level is -27 dBm. *(Pertains to examining a mixer output in a SA)*

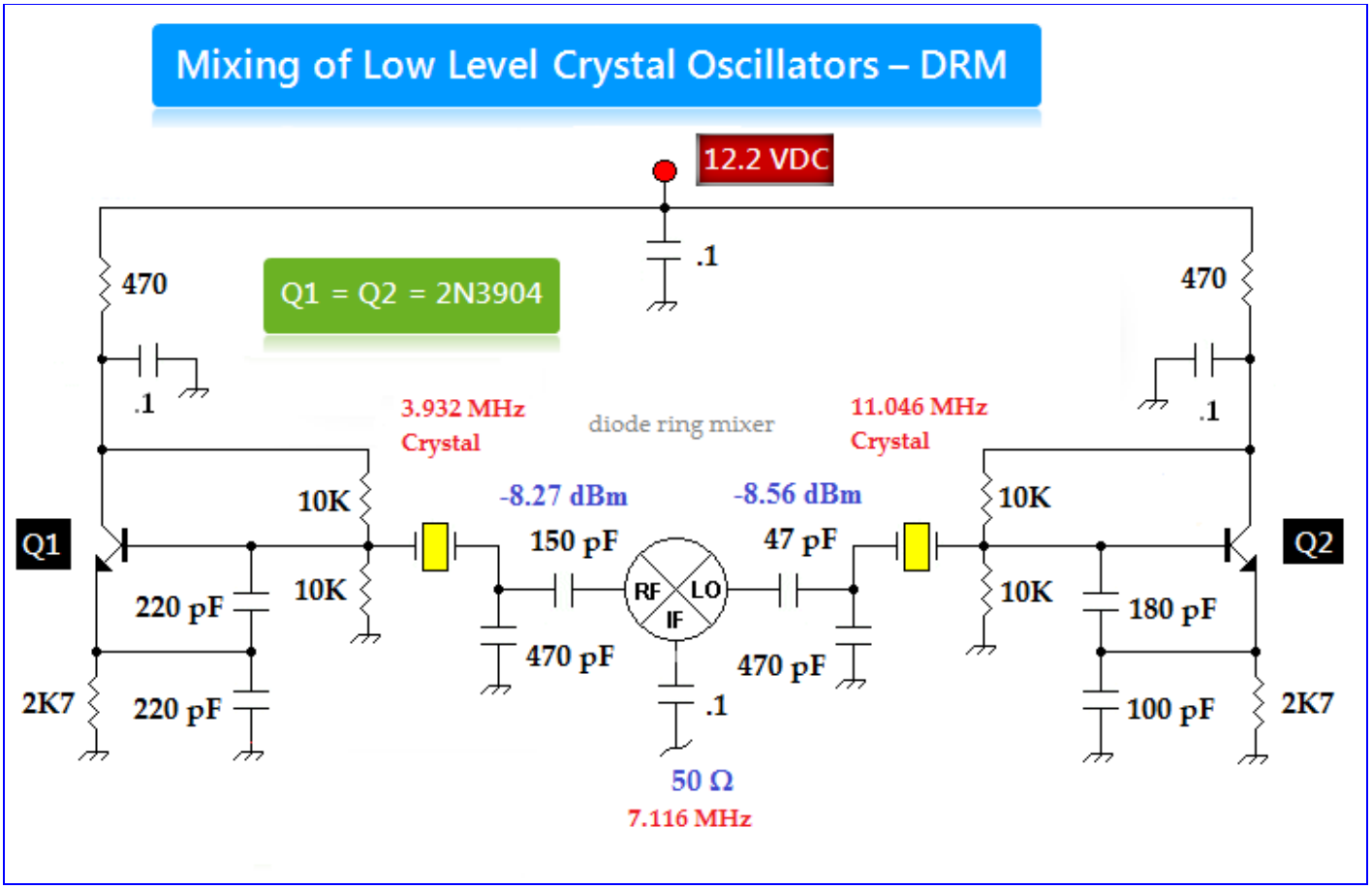
Choosing a Mixer

A number of mixers were considered; passive, active, unbalanced, single-balanced and finally, double-balanced. The diode ring mixer is an obvious good choice commensurate with my goals of reducing spurs, LO feed through and achieving high port-to-port isolation. In future web pages, other mixers may be presented, however this page is focused on the diode ring mixer. Click for a file with a few scanned pages concerning mixers from my "ideas only" notebook from ~2002. I own over 30 notebooks now.



Above — ADE-1 diode ring mixers. We're using these now as they're cheaper than the SBL-1, TUFF-1 etc. hole-though versions. Although SMT parts, they can be flipped over and wired "normally" with a little effort, steady hands and good vision. Mini-Circuits will sell them in small quantities to Hams; email them and enquire. I feel the diode ring mixer has been misunderstood by some amateur builders — lore and misperceptions that the 7 dBm LO port drive, the need for 50 ohm port terminations, a ~ 5 dB insertion loss and cost make them undesirable. Their excellent performance and design challenges are reasons why we use them; "the journey — not the destination" stuff.

In receiver applications, some builders and kit sellers seem more focused on features such low-battery indicators, digital displays, miniaturization and cost containment than basic receiver performance. Certainly keeping cost down down deserves consideration, however, good mixer performance is king. You'll have to decide what's affordable and important and build accordingly.

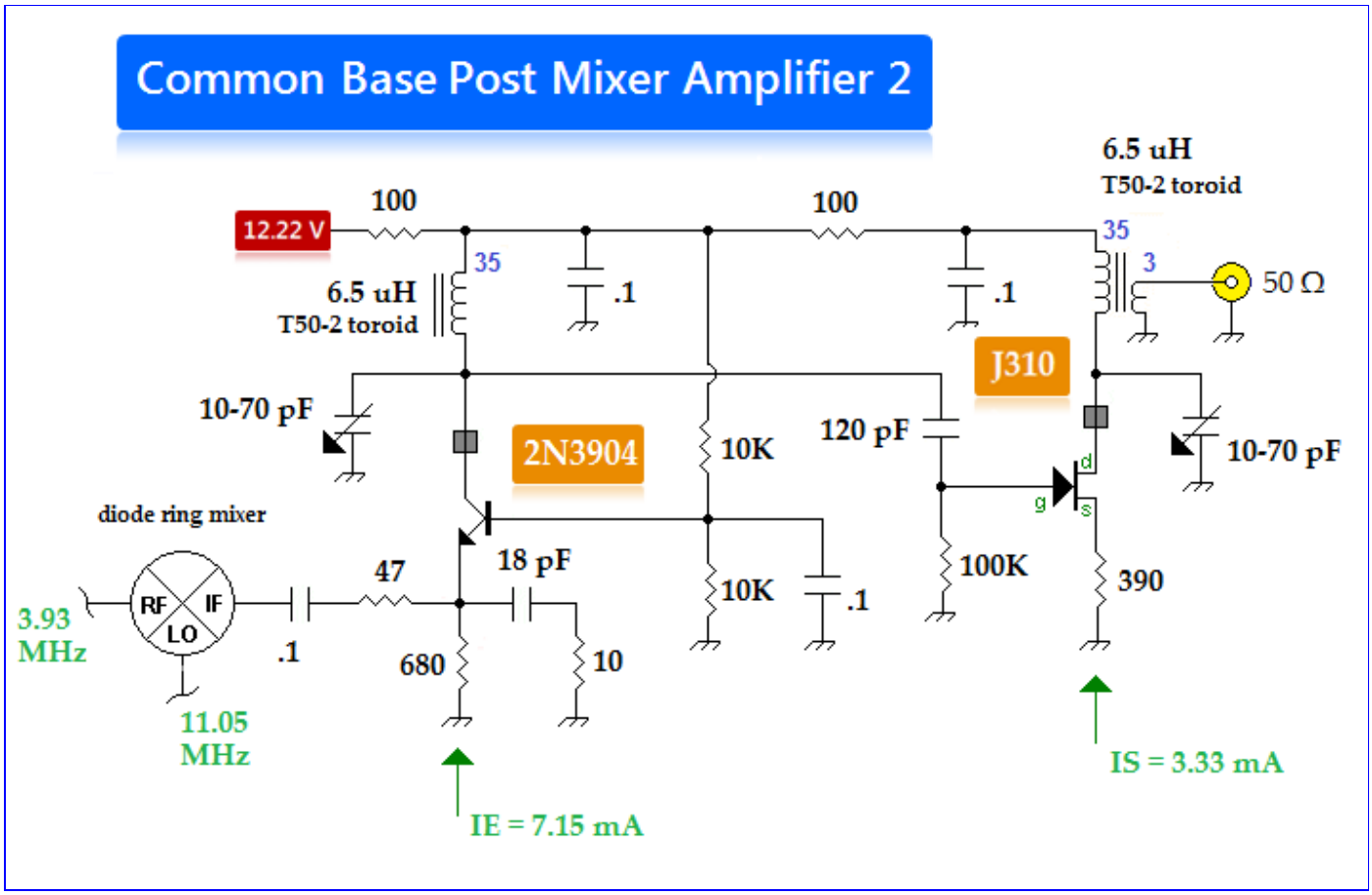


Above — My very first transmit mixer experiment. My hope was to build a transmit mixer possessing low spurious output to alleviate the need for stiff, post mixer band-pass filtering such as a triple tuned band-pass filter. Thus, low-level, low distortion output was taken from between each

crystal oscillators' shunt capacitor and crystal. The desired output frequency is ~7.114 MHz to build a transmitter for the 40 Meter Ham band.

The mixer output to 50 MHz looked like this in a Spectrum Analyzer. The dominant frequencies are the **sum** and the **difference**: 11.046 **+** 3.932 MHz = **14.98 MHz**; 11.046 **-** 3.932 MHz = **7.114 MHz**. The frequencies realized are *slightly* different since the oscillator output is shifted by crystal variances and from circuit capacitance.

In the experiments that follow, I built some circuits to filter and/or amplify the output of the *above* mixer circuit.



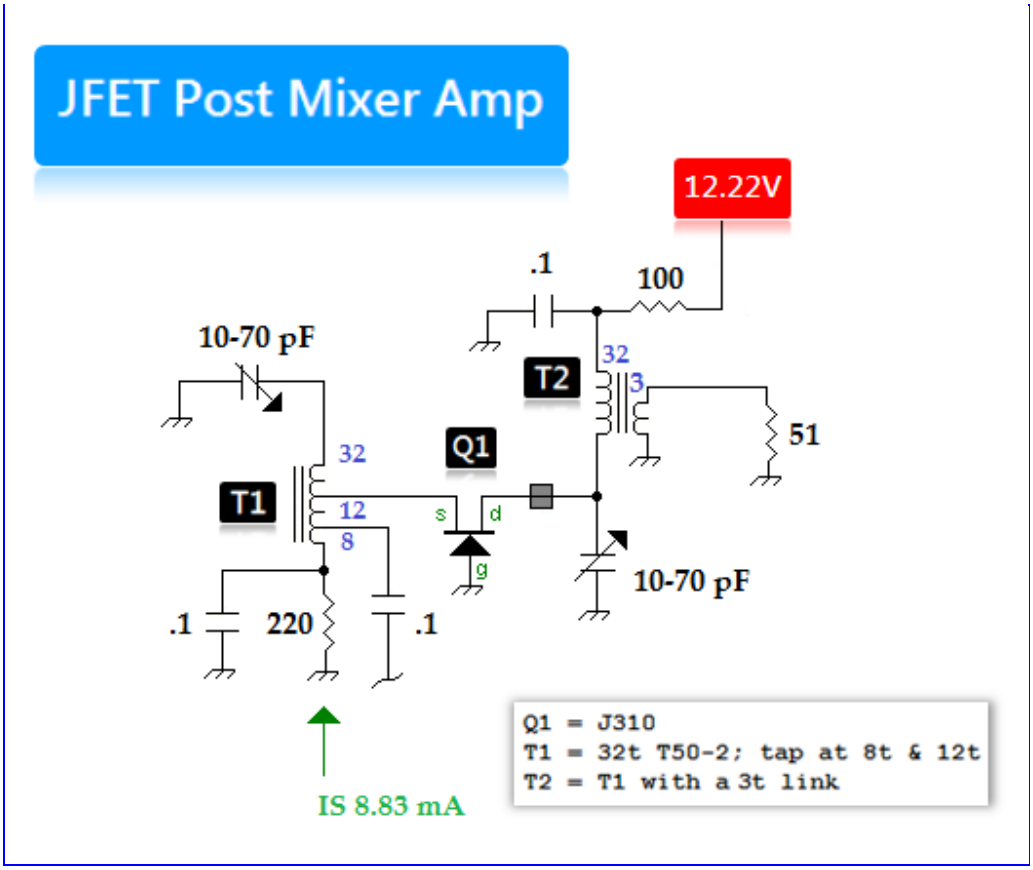
Above — The first post mixer amp; a common base input amplifier that's AC coupled to a common drain FET amp. I hoped that 2 tuned L-C tank circuits could substitute for a passive double or triple-tuned band-pass filter, plus provide some gain.

A broad-band, common base input amp was chosen to properly terminate the diode ring mixer and alleviate the need for a diplexer. A ~50 Ω input impedance is established by a 47 Ω series resistor since the 2N3904s input impedance is quite low due to the moderately high emitter current employed to boost gain and IMD performance.

This amplifier failed to reduce spurious output 50 dB down or greater — my design goal. Here are its scope and spectrum analyzer outputs; please observe that the unwanted 14.98 MHz signal is only 32 dB down from the desire IF of 7.114 MHz. An RC network consisting of a shunt 10 ohm resistor + an 18 pF cap provides additional low-pass filtering above 20 MHz. I attribute this simple filter to Dr. Ulrich Rohde as I have seen it in some of his post mixer, common base RF amplifier designs. Click for a brief supplement regarding his low-pass network ***(#2 RC Low-pass Network on the Supplemental Page)***

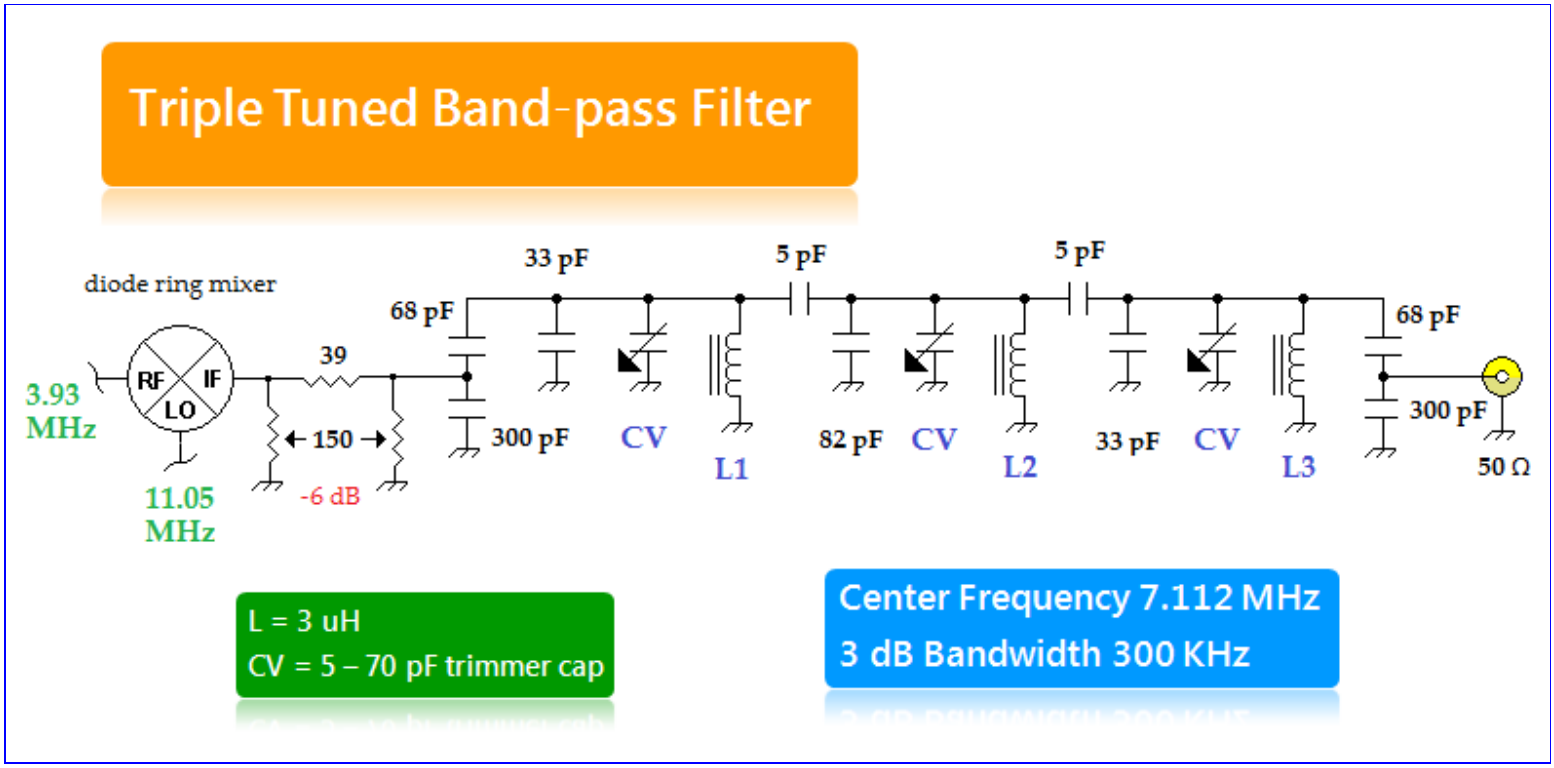
The amp design shown above was actually an improved version of this prototype. In the prototype amp, the mixer power at 14.98 MHz was only 23 dB down from the desired intermediate frequency of 7.114 MHz. You can't expect a single L-C tank to well filter a mixer output. Unfortunately, the 1K2 -12K resistor providing DC bias for the emitter follower lowers the Q of the common-base collector tank circuit. 12K is better than 1K2 in this regard. Poor performance sparked the design of the second common base amp shown above.

I use ferrite beads and 51 Ω resistors interchangeably on the collector/drains of amplifiers to snuff out UHF oscillations. According to my experiments, the resistors may work better. I purchased the ferrite beads from Diz. After the common base amplifiers shown above, I decided to try a tuned input + output common gate JFET amp:



Above — A JFET common gate amp with tuned input and output built about 3 years ago. For spectrum analysis. I padded the amplifier output to provide a -28 dBm 7.116 MHz signal. The vertical resolution on the SA is the standard 10 dB/division. As shown, the 14.98 MHz signal is ~ 39 dB down; an improvement over the amplifiers shown previously. This narrow-band amplifier requires a diplexer. I wanted *better* filtering than that offered by amplified circuits with 2 tuned L-C circuits, so I halted this experiment and decided to try a triple tuned band-pass filter.

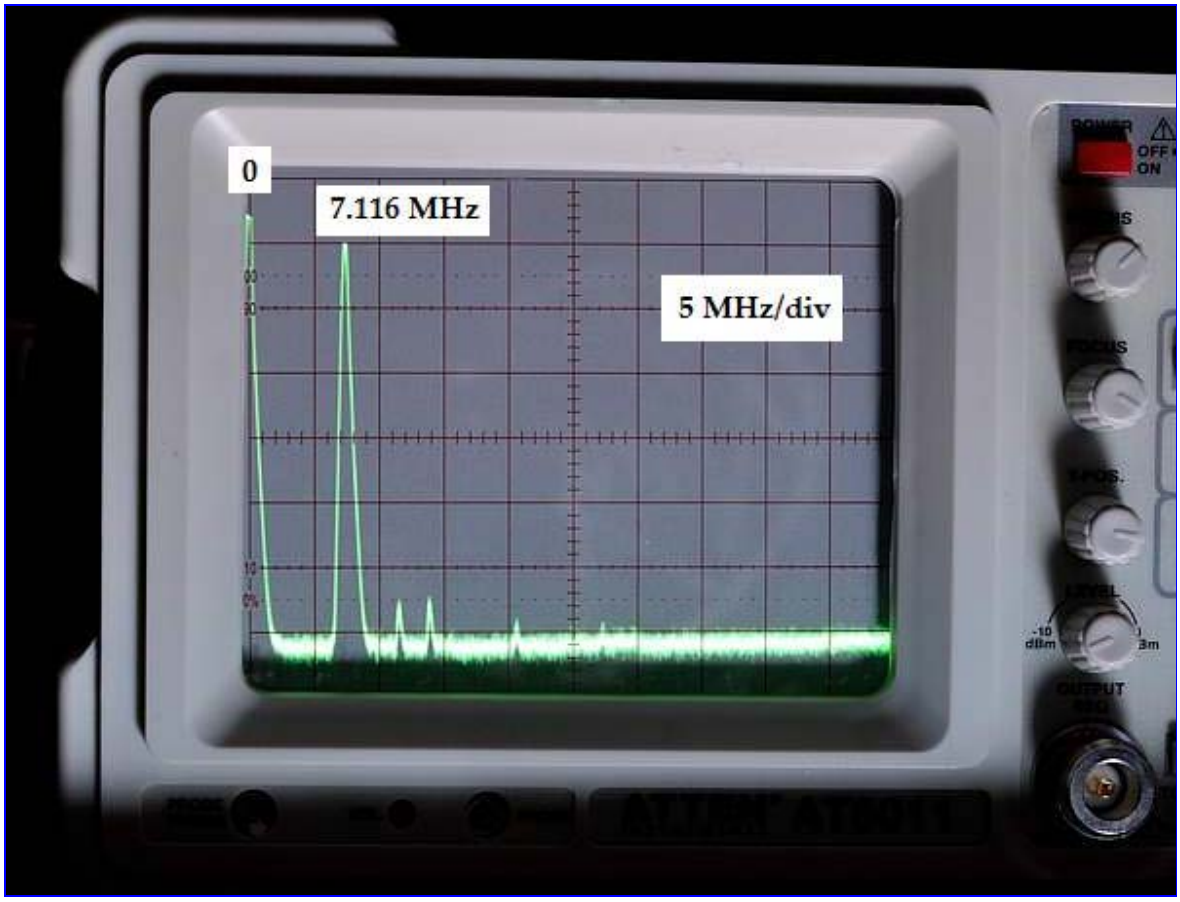
This JFET circuit experimentation spawned over 8 weeks of experiments concerning common-gate RF amps — some of them appear later.



Above — A triple tuned band-pass filter designed with software from EMRFD called TTC-08. Click for the breadboard photo. The diode ring was connected to the filter input via a 6 dB attenuator pad using short leads. Click for analysis in GPLA. All inductors are 3.0 uH — 23 or 24 turns on

a T50-2 powdered iron toroid. I measured all the inductors and obtain the exact desired inductance by expanding or squishing the wire turns, or if necessary, adding (this means rewinding the entire coil) or removing a turn. If you lack an inductance meter, just winding the formula calculated number of turns will be close enough for most applications — I only got a good L- C meter in 2009 and somehow managed.

I learned that the ultimate way to peak a triple tuned filter is by tweaking the tuning capacitors while it's connected to a spectrum analyzer — what a thrill!



Above — While a little tedious to build and align, the triple tuned filter worked magnificently; the strongest spur is 54-55 dB down and the 14.98 MHz signal is gone. Insertion loss = 2.5 dB. This experiment provided a benchmark of what great post mixer filtering looks like. Post mixer filtering is an important topic worth studying further:

Why do we need filtering on a mixer output?

Let's examine mixer ports more closely. A port is just a pair of wires where signals are applied or removed. There are 2 kinds of mixer *outputs*: 1) the sum plus difference frequencies; 2) spurs.

Further, 2 kinds of spurs occur: One type is straight feed through where 1 signal from the 2 input ports makes it out to a 3rd port. Examples include LO feed through to the RF port, or LO feed through to the IF port.

The other type of spur is a mixing product such as a harmonic.

In general, the mixer output frequencies are numerically described by an equation:
IF (output) = N x L +/- M x R
N and M are both integers, 1, 2, 3, ... L = local oscillator frequency, R = radio frequency

A mixer is said to be balanced when you duplicate some of its functions and then combine them — usually with transformers. Consider, for example, the single diode mixer — they work, but the output contains ++ feed through and spurs. A mixer with 2 diodes or 2 FETs etc. can be much easier to use because the transformer combines the signals in a way that cancels some of the spurs and feed through. The double-balanced diode ring mixer uses four diodes and 2 transformers — producing even less feed through and harmonic output.

In a double-balanced diode ring mixer, the LO and RF ports are balanced and all ports of the mixer are isolated from each other. The double-balanced mixer greatly reduces, but does not stop all LO feed through at the RF and IF ports. A wideband match at 50 ohms is required to maintain mixer balance; hence you will often see attenuator pads on the LO, RF and *especially* the IF ports.

Let's focus on the IF port. Attenuator pads absorb any reflected mixer products and signals coming back into the IF port, thus increasing the

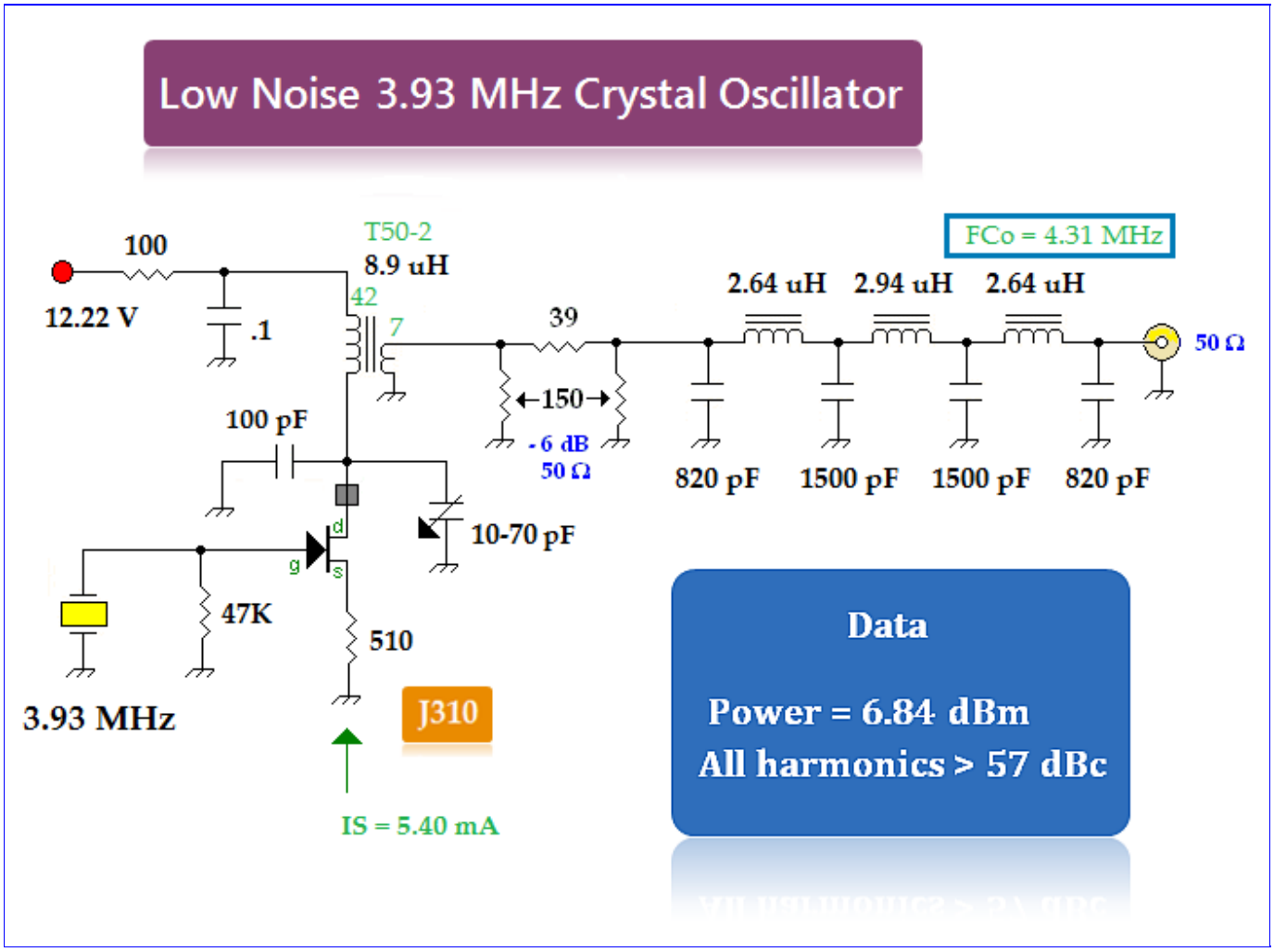
match to the IF port. You may have noticed some builders use a diplexer on the IF port. The diplexer presents a wideband match to all IF port frequencies — passing the desired sum or difference frequency and absorbing the unwanted mixer products reflected back into the IF port by subsequent stages.

Since the IF output contains the sum and difference of the LO + RF, LO feed though, and other spurious energy, band-pass filtering is required to launder the IF signal into something useful. Following a transmit mixer, we filter with an L-C band-pass filter — after a receive mixer, crystal band-pass filters dominate. If you choose an unbalanced mixer or single-balanced mixer, filtering becomes more difficult than with a double balanced mixer. Unbalanced mixers are usually reserved for situations where high performance is being sacrificed for cost containment and/or want of a low parts count. There is no free lunch — you either alleviate as many mixer products as you can at a low-level with good practices, or have to deal with them down your signal chain while sacrificing optimal mixer performance.

Double balanced mixers are sensitive to non-resistive IF port terminations. When improperly terminated, the 2 transmission-line transformers work poorly — any reflected power generates high voltage across the diodes and degrades mixer performance. According to Dr. Ulrich Rohde, some proper ways to terminate the mixer include using a diplexer followed by a wideband 50 Ω feedback amp, or a common-gate JFET amplifier. *(Reference 1)*

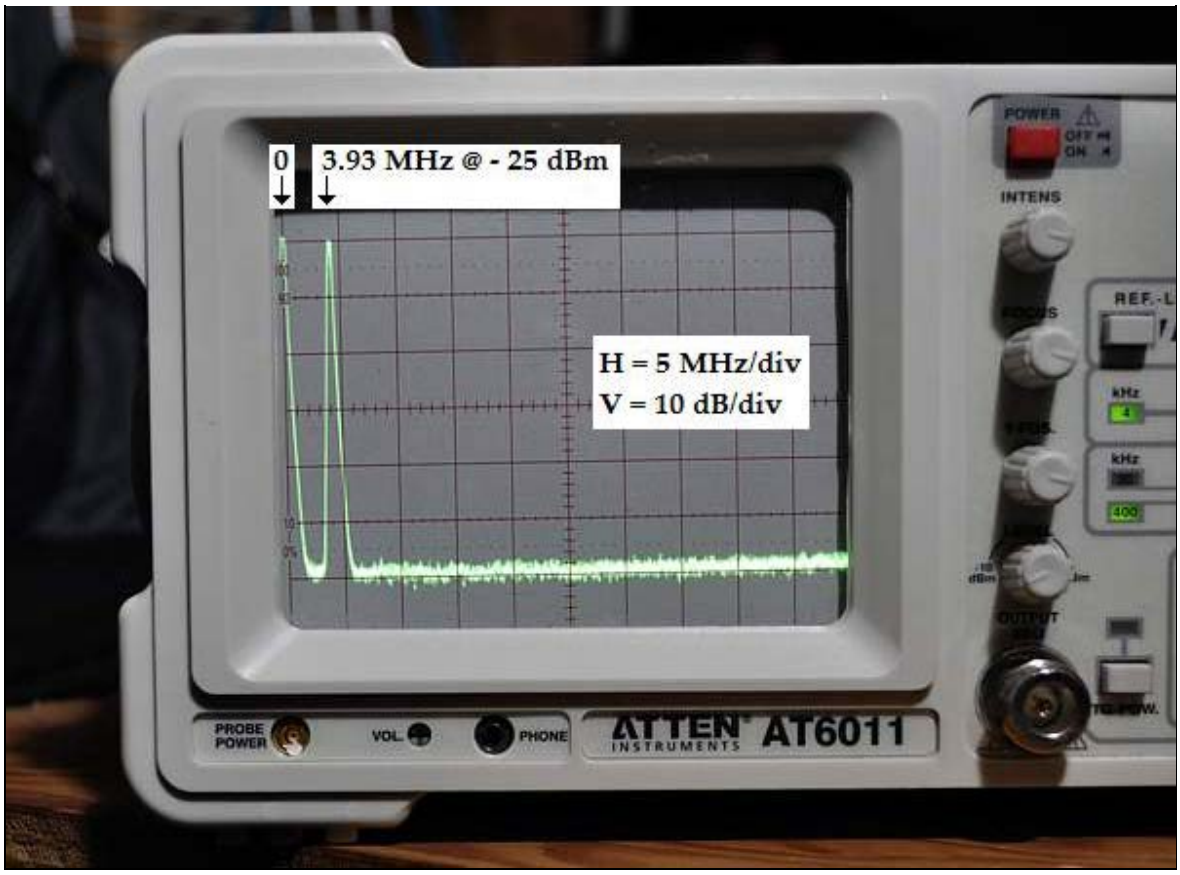
Improved Local and RF Oscillators

In the earlier experiment, I really should have the run the LO port at 7 dBm. In order to improve my experiments, new crystal oscillators were designed with emphasis on correct LO output power and low harmonic energy.

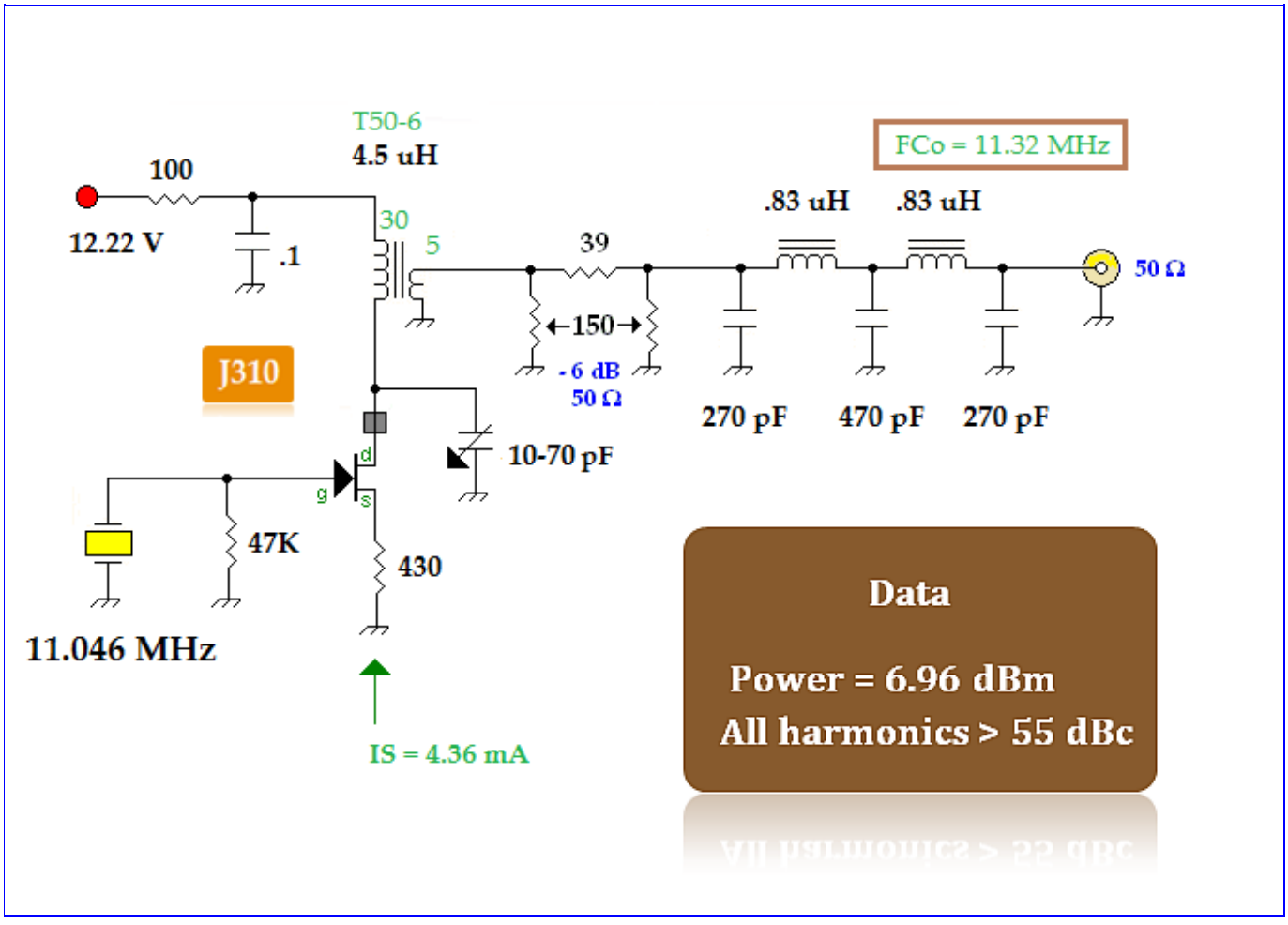


Above — The new LO; a 3.93 MHz crystal oscillator with stiff low-pass filtering. Admittedly, this 7 element Chebychev low-pass filter is overkill, however, I wanted to examine filtering and learn how much is required. On the bench — do whatever you like; even chasing crazy personal goals can be instructive and help you relate to information from texts and articles, or satisfy a whim. I can read something 100 times, but may not understand it well until I actually do it.

Above —The LO breadboard in close-up using a long focal length lens. Click for a wide angle photograph. The unsoldered end of the 100 Ω resistor in the close up photograph is where I connected the VCC.



Above — Spectrum analysis of the well-filtered 3.93 MHz LO. Not surprising, no harmonic energy is seen.

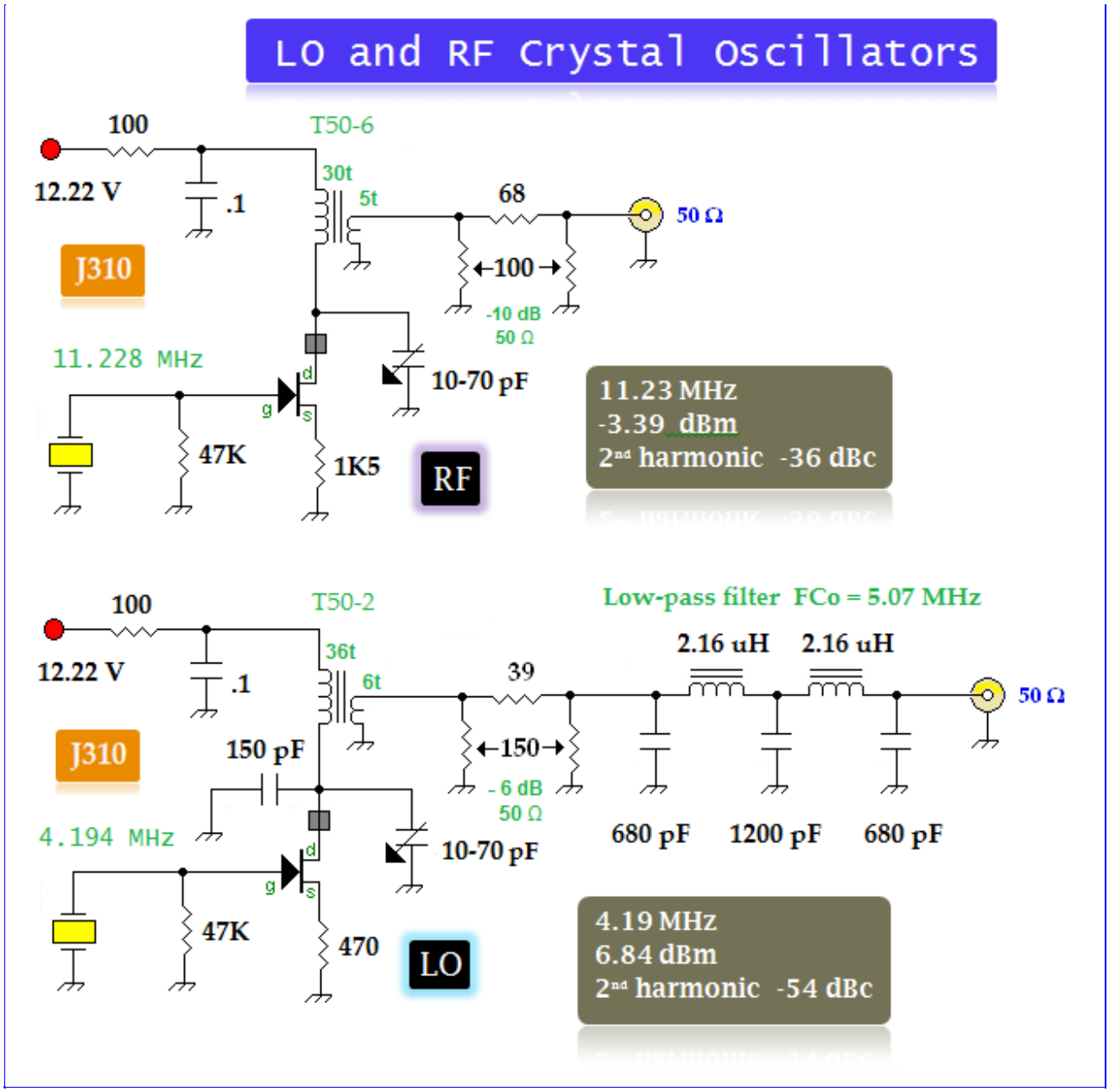


Above — The redesigned RF port oscillator. Clearly, the 7 element Chebychev low-pass filter isn't needed, so an N = 5 version was tried. Click for the spectrum analysis — again no harmonic energy was seen. In both the RF and LO signal generators, I tried to get as close to 7 dBm output power as possible.

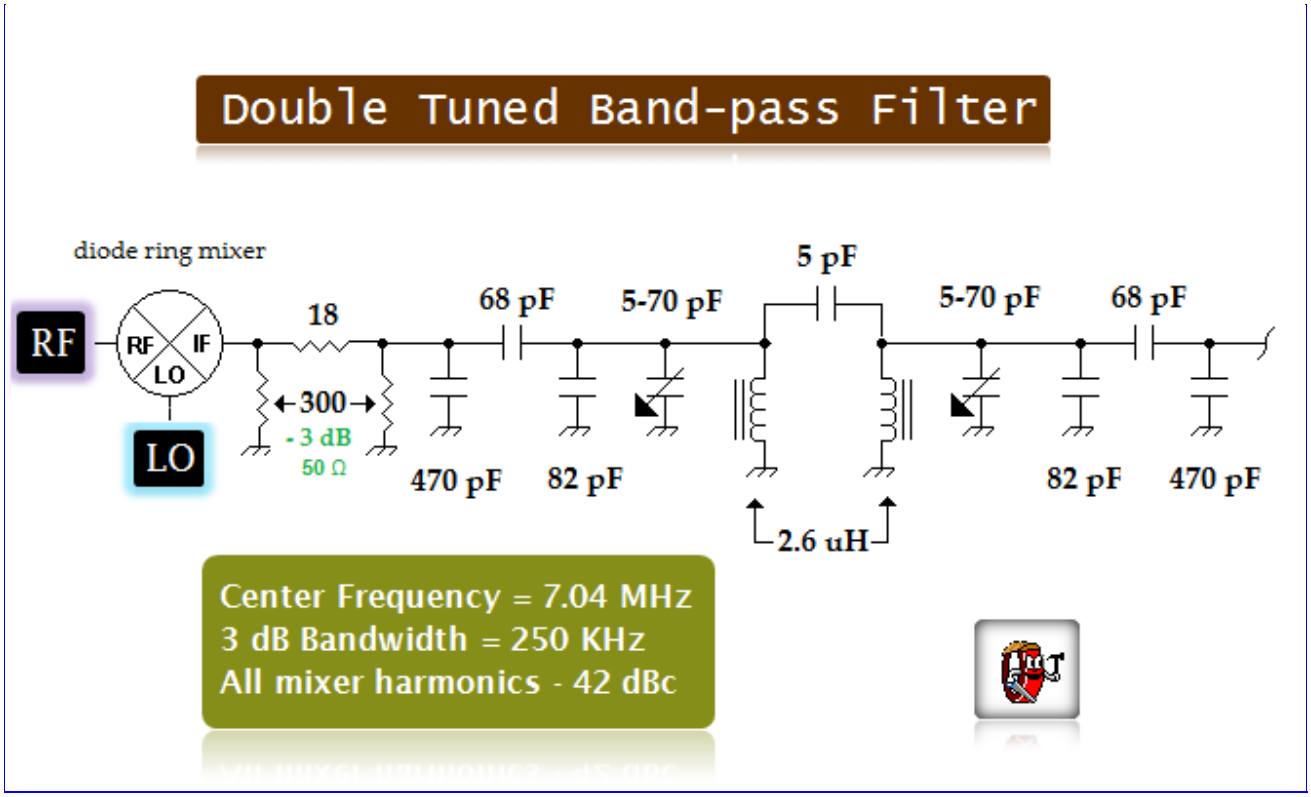
To operate this oscillator at the desired RF port signal level; for example, between 0 and -10 dBm, you might just attenuate the output with a fixed pad or step attenuator. My conclusions echo the work of others experimenters; lowering the RF port down from 7 dBm to as low as - 10 dBm, lowered the amplitude of the spurious mixer products seen in the spectrum analyzer. Click for a sample. Refer to the QST Technical Correspondence citation in the references section for more information.

Above — The 2 re-designed oscillators wired up and connected to an SBL-1 mixer. I actually connected the attenuator pads after the low-pass filters as explained later.

Two different crystal oscillators were then built:



Above — Two different crystal oscillators targeting ~ 7.034 MHz were built. Click for the breadboard photo. You can see the crystal frequencies in this photograph. The RF port oscillator power was set to -3.39 dBm by choosing a low value JFET source resistor and attaching a 10 dB attenuator pad. Relatively low harmonic distortion prompted the exclusion of a low-pass filter on the RF oscillator. The LO output power was ~ 7 dBm.



Above — The final experiment; placing a double tuned band-pass filter after the TUFF-1 diode ring mixer with the 2 latest crystal oscillators attached . This filter was in my junk box and I peaked it for the 7.034 MHz IF with a spectrum analyzer. The strongest spur was 42 dB down from the carrier — falling well short of the triple tuned band-pass filter presented before.

Clearly from all these experiments, a strong case for placing a triple tuned band-pass filter after a transmit mixer exists. If you use an unbalanced or single-balance mixer, a double balanced mixer might sufficiently not block feed through and spurious RF to keep your signal chain tidy. I enjoy studying the transmit chains of others to see how they filtered spurious and feed through RF. At the end of the day, as long as the output carrier spurs are low enough to meet your country's regulatory requirements, you're okay. Designing for low spurious emissions is an exciting challenge — one you'll miss if you don't try your hand with RF design.

A realization emerged following these experiments — I couldn't measure the return loss of the local oscillators! It technically could be done, but not by me. After 2 weeks of struggling, I engaged an American colleague with whom I occasionally build experimental circuits. After making some progress, we became stalled again. This time I asked Professor Ken Kuhn and Wes, W7ZOI for some ideas. Eventually a method to measure the RL of local oscillators came together along with enough material for another web page — RF Workbench 3.

When you do experiments, knowledge evolves as you go — for me, I learn mostly from making mistakes. I often think I should repeat most of my experiments over before presenting them, but this would consume too much time. However, footnotes can serve to steer readers for minor issues. If I had to re-build the crystal oscillators from Part 1, I'd build each crystal oscillator with a separate JFET buffer — then the return loss of the oscillator buffers could be measured as shown on RF Workbench 3 (with a shorted oscillator tank). Also, the pi attenuator pads on the crystal oscillators should follow the low-pass filters to garner the best output return loss. The good news is past experiments inform future experiments.

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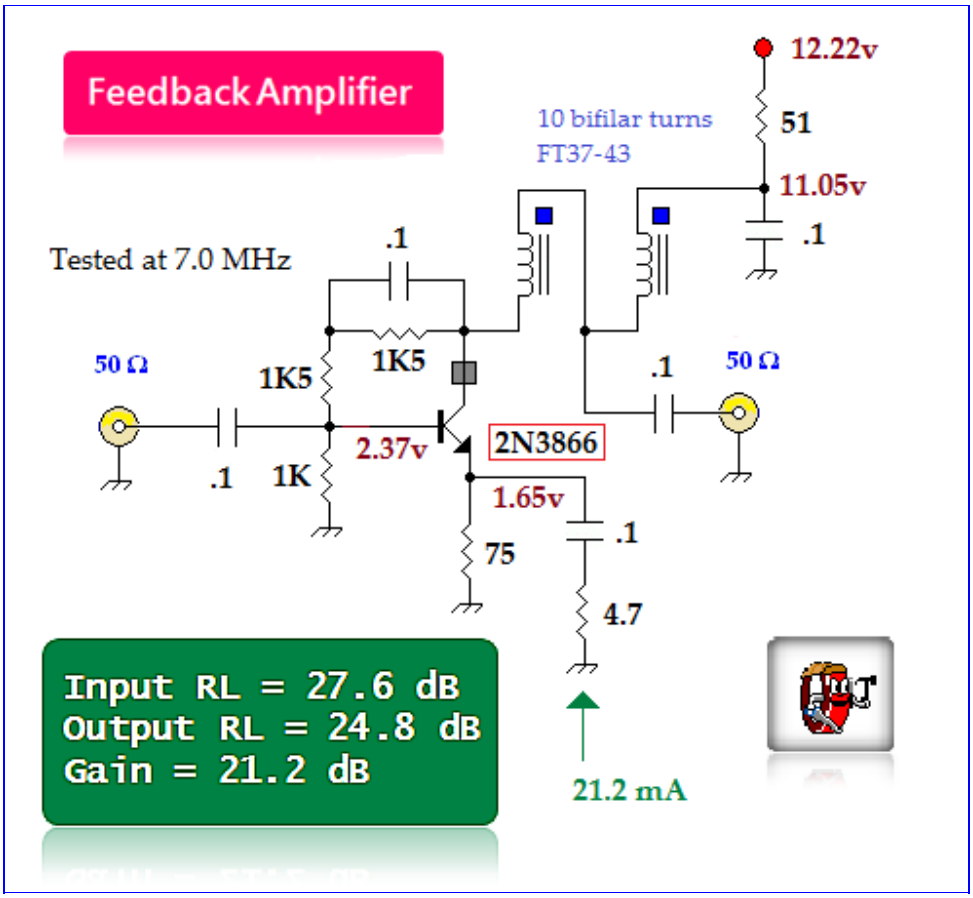
2. Bipolar Transistor Feedback Amplifier Experiments

I love making signals bigger — especially while preserving fidelity. It would be nice to become a reasonably competent amplifier designer — hopefully by studying sound schematics, applying software, building circuits and measuring evermore parameters this might occur. The mathematical equations of RF amplifier design seem quite daunting; they're the fodder of electrical engineers with their Hewlett Packard scientific calculators, SPICE software and GHz F-t transistors. With most things technical, as you try to advance, more questions than answers cross your mind; however, somehow this is normal and may actually signal progress.

Abundant amplifier references exist; for example, EMRDF Chapter 2: Feedback Amplifiers. This is essential reading and I won't repeat this information. Rather, I'll just share some ideas developed or reinforced on my bench. In the past, I've preferred amplifiers with narrow-band (tuned circuits) in an attempt to reduce distortion and maximize gain. Now after critically examining these tuned amps with scope and spectrum analyzer, I better appreciate the significant intrinsic feedback of RF transistors ***(the tendency to oscillate)*** and broadband designs are sought. Often you'll spend more time taming a tuned amplifier than building one.

This section focuses on return loss, bias techniques and achieving linear amplification — for example; finding ways to apply negative feedback,

match the input, or how to set the collector voltage. All my experiments and thoughts about RF amplifiers are from an amateur designers' perspective and I welcome your feedback. The first amplifier shown is a classic W7ZOI topology that I call the "Beaverton Special".



Above — A classic feedback amplifier popularized by Wes, W7ZOI in books like Solid State Design for the Radio Amateur and EMRFD. My respect for this humble design increased after building and testing 4 different versions to get a feel for amps with both shunt and series feedback. Of the 4 built, this particular version became my favorite — providing excellent input and output matching without crazy high emitter current. Employing a low noise / high F-t 2N3866 transistor is icing on the cake — an attempt to maximize impedance matching and performance using this standard, fits most transistors bias/feedback circuit. The humble 2N3904 also worked well in this slot. You don't need the ferrite collector bead with a 2N3904.

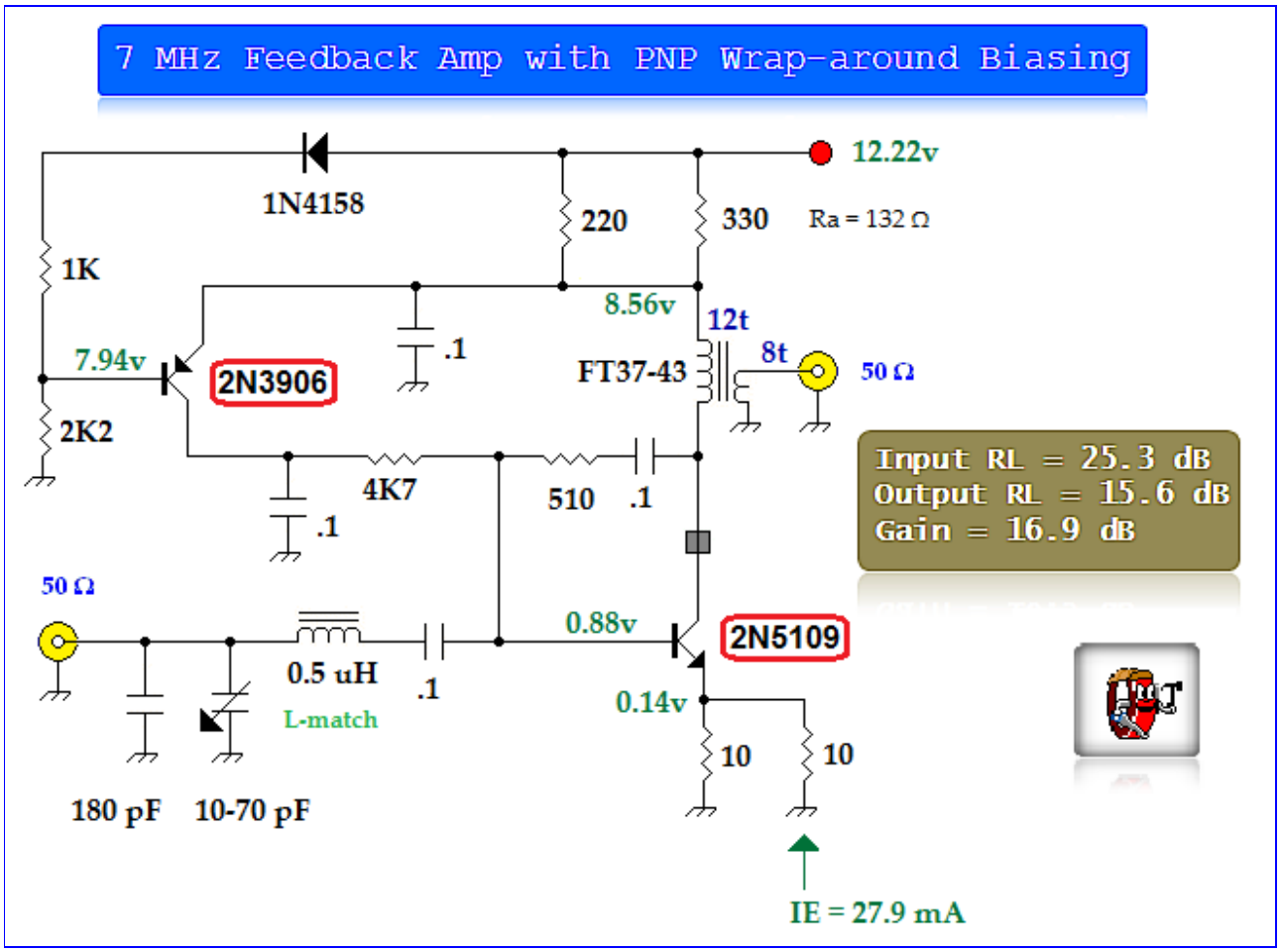
Other good experiments include trying different transistors and/or increasing the emitter current while being careful not to exceed the BJT's current rating (plus add heat sinking as required). You might also try the stage at different frequencies or perhaps sweep it to see at what frequency the gain starts to fall off.

Of the 4 BJT feedback amps shown in part 2, only this amp has a true broad-band input and output. What bothers me about broad-band linear amplifiers is that when you chain up 2 or more of them, signal fidelity generally degrades as it passes through each successive amp stage. Solutions include mopping things up with some low-pass filtering after the last stage, leaving it alone, or tuning the amplifiers (i.e. not using broadband stages).

The biggest caveat for feedback amps are variations in input and output impedance caused by source and load mismatches. For example, a 75 Ω resistor was connected to the output of the FBA above. The input return loss degraded to 16.8 dB. Further, the same 75 Ω load was removed and then connected to the input during output return loss measurement— this degraded the output RL to 23.4 dB. Clearly *load* mismatches upset return loss more than *source* mismatching. A 50 Ω attenuation pad should likely follow a feedback amp in situations where high input return loss are desired; for example, after a diode ring mixer.

Noticing a variation of the classis feedback topology in EMRFD Figure 6.140, I asked Wes, W7ZOI about it. It turns out there's another way to "skin the shunt feedback cat". The above RF amp uses a series connection of 2 feedback resistors (1K5 and 1K5 with a bypass cap across one 1K5). The result is a resistance at DC of 3K, but a resistance at RF of just 1K5. You could also use a 3K resistor directly from collector to base that is paralleled by a series connection of a 1K5 resistor plus a 0.1 uF capacitor. That network has the same impedance as my amp shown above. That is; the resistance would be 3K at DC, but 1K5 at RF.

This explanation fueled the next experiment — transistor amplifiers have 2 operating conditions; 1 at DC, the other at AC. Like a carpenter framing a house, you begin design by setting the DC bias — no small design task since bias concerns more than just establishing the base voltage and emitter current. For example, biasing may effect voltage gain, maximum signal handling capability, noise figure, impedance matching, class of operation, the operating point (sometimes called quiescent point or q-point), feedback and temperature stability. Biasing provides much to think about, however, a practical way to explore any topic is to chunk it into small, understandable pieces that become a stepping stone to advancement. Let's focus on biasing for temperature stability. The next amp uses the wrap-around PNP bias — an awesome technique.



Above — A 7 MHz FBA using PNP wrap-around biasing. I learned about wrap-around biasing from Wes, W7ZOI and share a simple way for new builders to also learn this technique as the **#1 Design Center** on the supplemental web page. Click for a prototype breadboard photograph. This amplifier employs heavy shunt feedback from collector to base. Degenerative (series) feedback from the 2 parallel 10 ohm resistors also enhances temperature stability.
Expanded bias circuit temperature stability discussion follows amplifier number 4.

The wrap-around or feedback bias scheme is good because it's self stabilizing. The diode in the PNP bias network further ensures that the PNP bias remains constant with temperature changes. It really should be to glued to the NPN transistor (or its heat sink) to allow tracking of the NPN's

temperature variations. This bias circuit doesn't load the NPN base input impedance. Another great virtue is that the emitter of the amplifying transistor can be directly connected to ground allowing better performance at VHF and UHF.

Noise from the PNP will be amplified by the NPN, so the low-pass network formed by the 0.1 uF capacitor and 4K7 resistor is essential. In some related circuits, you may see an RF choke used instead of a resistor.

The actual 2N5109 input impedance is probably around 40 ohms — easy matching with an L-match network.

Amplifiers with an L-match tuned input shouldn't follow a diode ring mixer unless preceded with a diplexer since the narrow-band L-match tunes only 1 frequency. L-match networks can make an impedance bigger or smaller depending how they're oriented and also provide some low or high-pass filtering depending on the configuration. I design my L-networks on the bench using experience plus trial and error — a better way is to use software. I recommend the program called Zmat08.exe that is included on the CD that accompanies EMRFD. The software will get you close, however, bench tweaking is required since you're often matching a complex impedance, comprised in part, of stray reactance.

Setting up an L-Network for an Input Match

A suggested bench method for optimizing input Return Loss (RL) using an L-Match network.

Your task is find the "perfect" L and C values to get a RL of 20 dB or higher. Start by soldering in an inductor calculated from Zmat08.exe or according to your wisdom. Set up the amplifier for input return loss measurement. The first chore is to find the nominal target capacitance that provides the best match at the design frequency. I use a big range, air variable capacitor for this — with the input circuit connected to a return loss bridge, connect up and tune the big variable capacitor to give the greatest RL. Remove the variable capacitor, measure it, and then solder in an equivalent trimmer capacitor, and as required, fixed capacitor(s) so you can tune at least 25 pF above and below the target capacitance. Often, the target C will be close to whatever the software recommends. In amplifier 2, my C values are the 180 pF + a 10-70 pF trimmer.

Next, determine the optimal inductor. On my bench, I keep a variety of pre-wound #6; and #2 material powdered-iron toroid inductors and choose one close to the calculated or a self-chosen L value. I start with an inductor wound with 4-5 more turns than needed. After soldering it in, the RL is checked. Remove 1 or 2 turns, tweak the trimmer capacitor and again check the RL. If after removing 1 or 2 turns, the RL is going up, you've determined there was enough inductance to get the best RL. *(If the RL goes down, you probably didn't start with enough L to get the best possible RL).*

You can also also squeeze together or spread apart the toroid windings to vary inductance — the maximal inductance variation varies due to factors including wire gauge and total turns. Compressing the windings with thumb and forefinger *increases* the inductance and widening the gaps between windings *reduces* inductance on a toroid.

Assuming the RL increased after removing 1 or 2 turns, remove another turn, tweak the trimmer capacitor and check the return loss, and so on. Repeat until your return loss starts to decrease. Then add back a turn or 2 to find the absolute best match. *This procedure allows you to find the optimum inductance in-situ.* Once, you've figured out the best inductance, cut the inductor leads short, solder it in, tweak the trimmer capacitor, and then consider further tweaking the coil by expanding or squishing the windings on the toroid while looking at the RL in a bridge detector.

In summary, to get the best possible RL — design a prospective L-match with software, and then bench test to determine the optimal in-situ L and C by using values above and below the calculated L and C values while observing the results in a return loss bridge. This method seems tedious, but emphasizes that repeated bench practice and patience pays off. You can always just use the calculated L-network values and/or develop your own method to set them up.

Consider mitigating the stray inductance caused by the long lead that occurs after removing wire turns by cutting the lead and scraping off the enamel insulation every couple of turns or so. This is a gamble — If you cut the lead and need to add back a turn, you'll have to rewind the coil from scratch, or add in and solder another turn (messy). I'm often able to get a L-network RL of 22-26 dB using my method and feel it's worth the the time and effort.

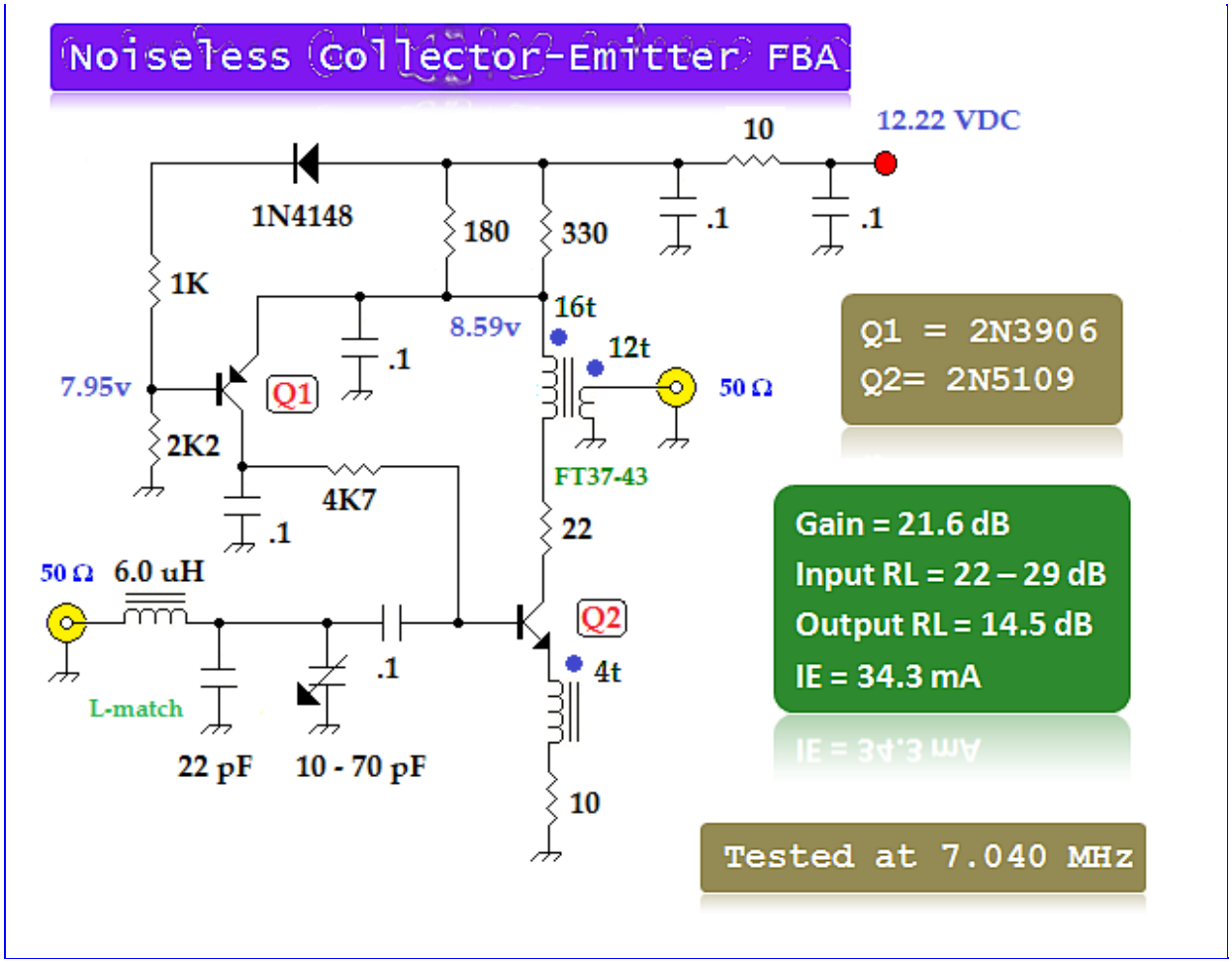
When bench tweaking the L and C values, your actually looking at the peak-to-peak AC voltage with the amp input connected to the unknown port of the RL bridge. Tune for the lowest, stable peak-to-peak voltage. Test it against the open circuit peak-to-peak voltage to calculate the RL. Since the open circuit doesn't change, you know the return loss is improving when the peak-to-peak voltage of the amplifier under test is going *down*. I store the open circuit voltage in my scientific calculator and calculate the RL from time to time as I'm tweaking the L and C values. After awhile, the whole procedure becomes automatic and quick. Once you 're done and everything's tidy, measure the open circuit and connected amp peak-to-peak voltages and calculate RL a final time. This is your reportable return loss.

You can scale matching networks from other builder's schematics by calculating the XL and XL and then applying these reactances to your desired frequency. Bench tweaking is still required. I also hope the person whom I'm copying didn't make a bench or drafting error. Be discerning about whatever your find on the Internet "Misinformation Highway" — this site included. Although I'm no philosopher, I know at least 3 things about people. They: 1. are often biased; 2. can lie; and 3. can make errors.

Above —Some toroids and the *air variable capacitor* I sometimes use to **coarsely** bench tune L-C circuits to determine the "ballpark" tuning capacitance. This capacitor features built-in reduction drive and varies from 15pF to 428 pF. When using an external capacitor connected to your circuit with short copper wires, expect some signal distortion and watch out for hand and body caused capacitance variations. The connecting wires also have reactance which won't be there when you swap in a small trimmer cap plus any fixed value capacitors.

Next up is a common emitter amp using "noiseless" feedback - this means the AC feedback is achieved with transformers instead of "noisy" resistors.

3



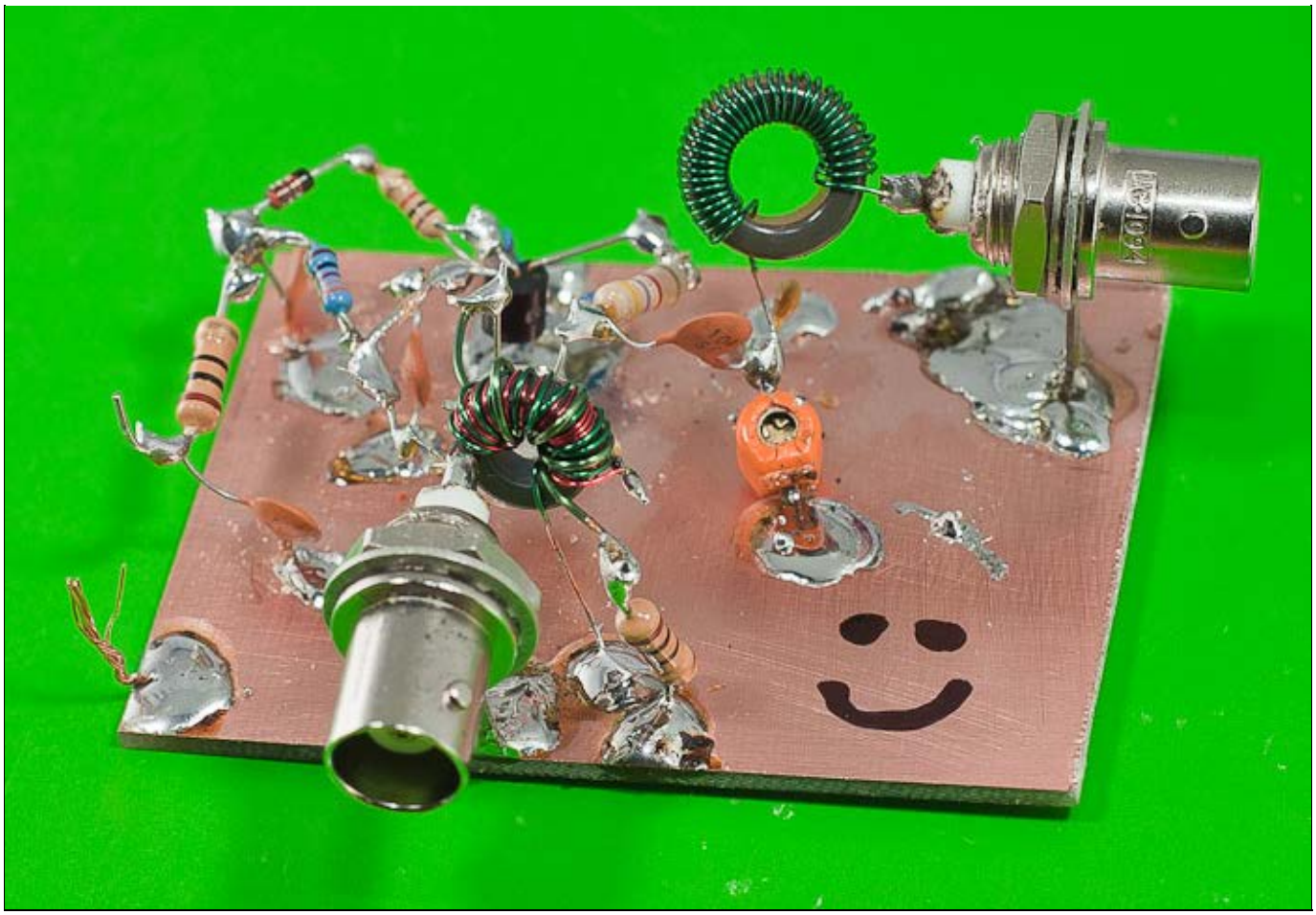
Above — Schematic of a 7 MHz collector-emitter "Griffiths" feedback amp. I ran substantial emitter current through this NPN. RC = 116.5 ohms — I paralleled 2 resistors for RC because I lack resistors between 100 and 150 ohms. The basic design is by Bruce Griffiths, who has a great web site. I thought I put up *big* schematics!

The input L-network was designed on the bench and provides a good input match peaked at 7.040 MHz — this pumped up the gain 3-4 dB. In my amp, a T50-2 powdered iron toroid inductor forms the L-match coil. Matching for the best possible input return loss is touchy and best done on the bench. For example, if the 6 uH inductor is decreased to 5.8 uH, the match could fall by 2-4 dB. With patience and careful tweaking return losses approaching 29 dB are possible, but likely too time consuming for most builders. The procedure as described earlier is pragmatic: connect a RL bridge to the input and adjust the L and C values until the lowest return loss is discovered. Even squishing or expanding the toroidal inductor windings can squeeze out a final dB or so of input matching.

Output matching proved interesting. Although I tried, the best output RL I could muster was 14.5 dB. Lowering the 10 Ω degeneration resistor or increasing the current could increase the output return loss. An output attenuator pad might be considered — a 3 - 6 dB pad would increase the output RL to over 20 dB.

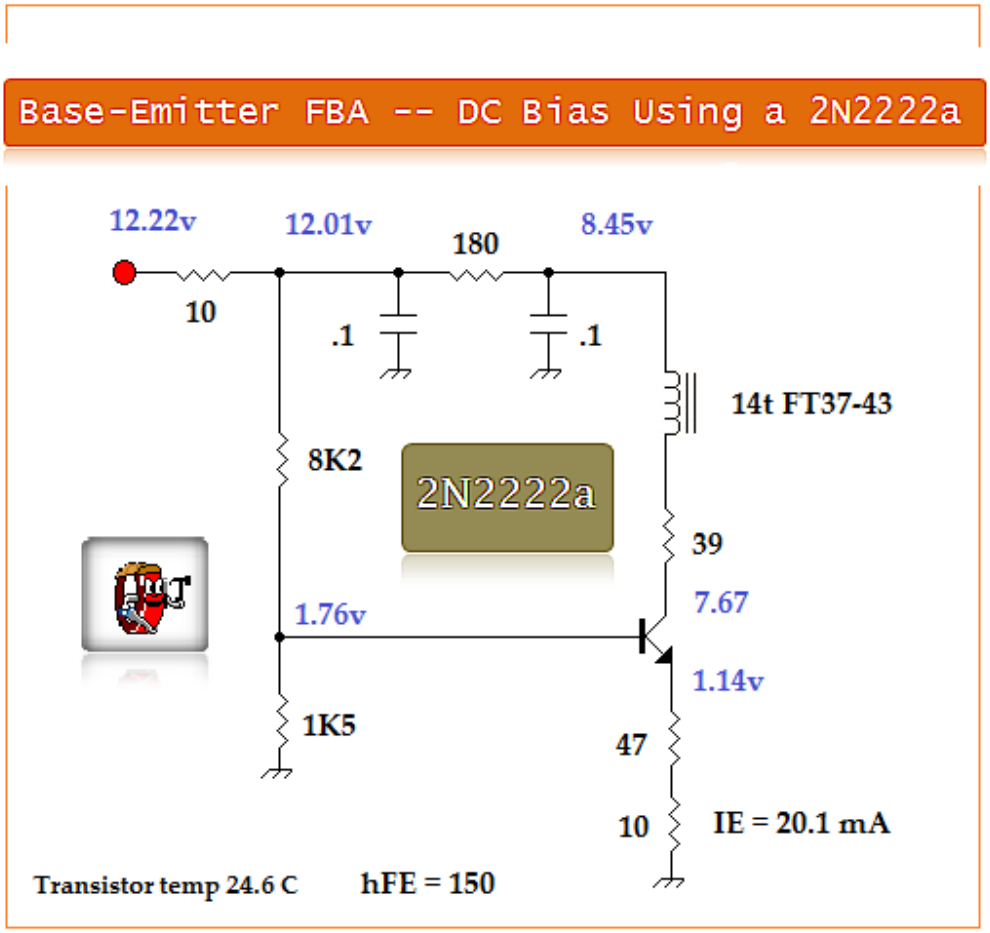
All 3 output transformer windings were wound on a FT37-43 with care to keep the phasing correct. Amplifier gain is not dependent on collector current. For example, substituting an Ra of 180 ohms (clipping out the 330 Ω resistor) yielded a gain of 19.5 dB, an emitter current of ~ 20 mA and an output return loss of 12.6 dB, while the input match changed very little.

The oscillation snuffer 22 Ω collector resistor was 15 ohms in another version, however, parasitic oscillations were discovered at ~175 MHz and snuffed out by raising this resistor from 15 to 22 ohms. I sometimes go as high as 51 Ω; especially in JFET circuits.

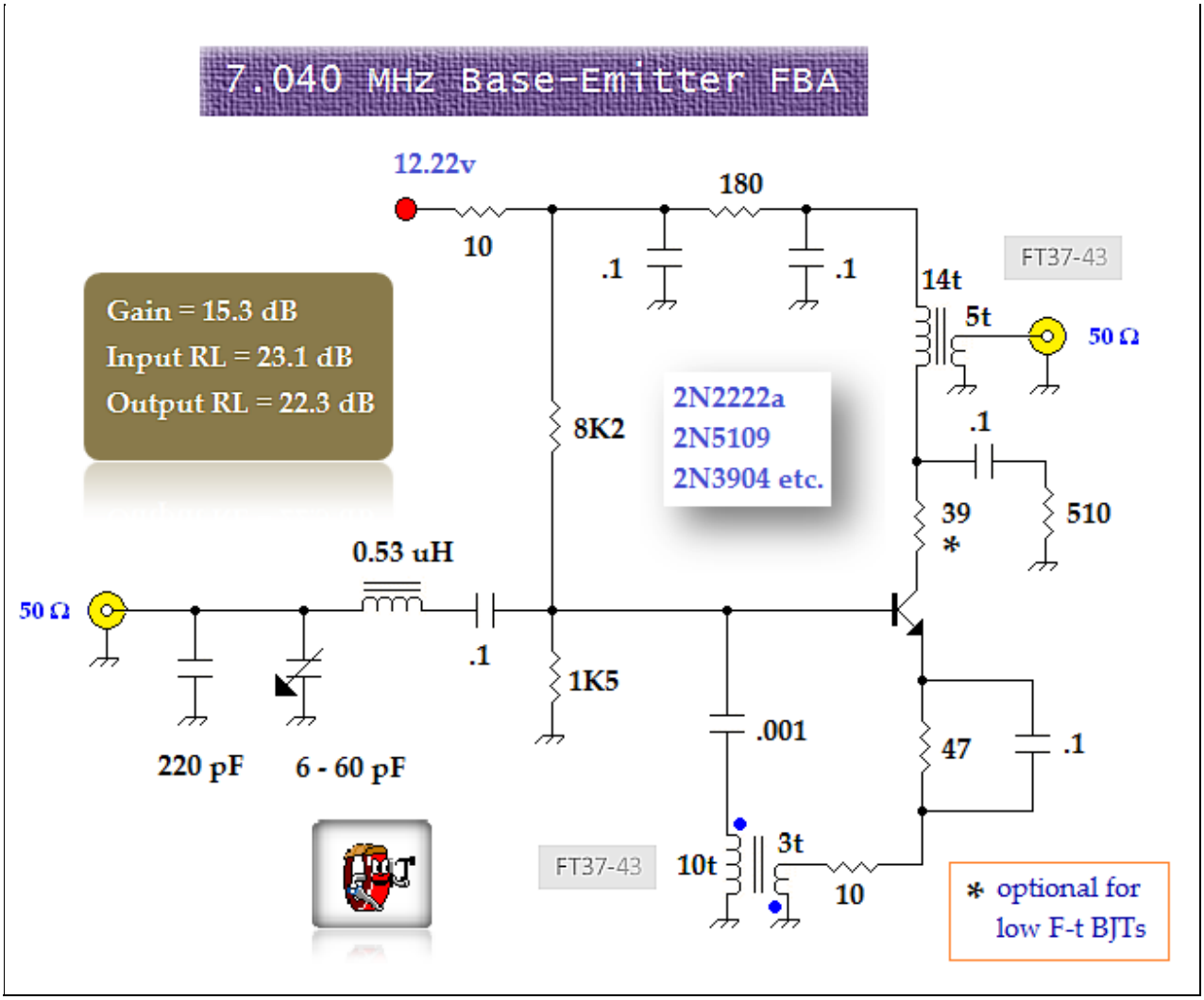


Above — The breadboard of the noiseless collector-emitter 7 MHz feedback amp. Click for a photo of another version. The hot "modern" replacement for the 2N5109 is this SMT part. I also like the BFG135 T/R BJT.

The final FBA experiments *below* use a standard voltage divider bias, tweaked for temperature compensation. The AC feedback is base to emitter — a rarely used topology in North America; although I'm not sure why.



Above — The DC bias resistor values for a 2N2222a with a DC Beta or hFE of 150 and a emitter current of 20.1 mA. Almost every text author writes about voltage divider bias temperature stability, but some builders get bogged down in the details. Since the bipolar junction transistor is a voltage controlled device *(see section 4: QRP-POSDATA for an explanation)*, you must set up some DC voltages — I created a design center presenting an easy approach to design reasonably temperature stable BJT amps. See **#5 Design Center** on the supplemental web page. After getting the bias, the AC parts were added, and the completed schematic is shown below.



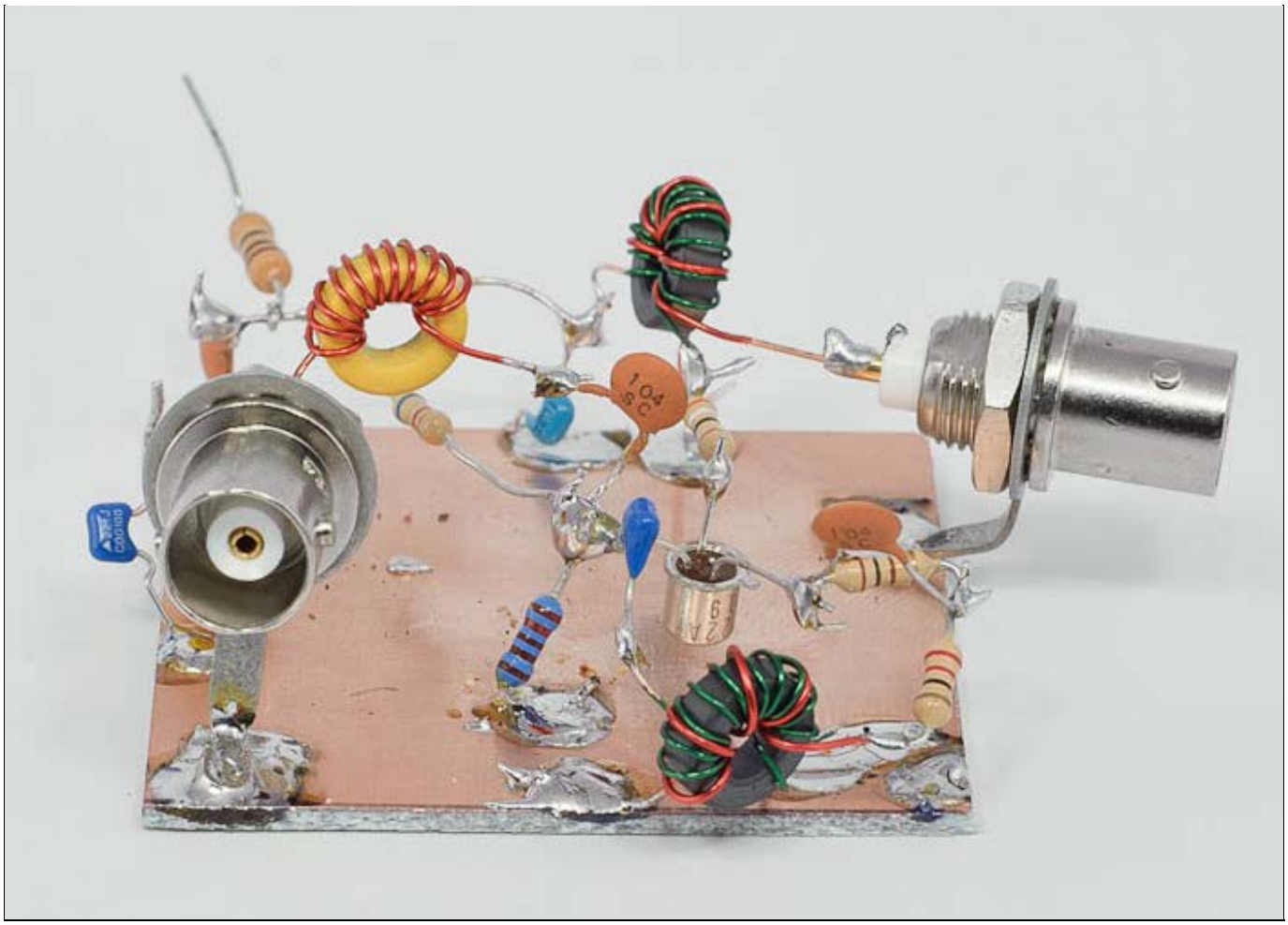
Above — A base-emitter feedback amp built Dec 21, 2010. I read about base-emitter feedback in Dr. Rohde's book *(Reference 1)*. He had some discussion, a small signal model and lots of difficult math, but no circuit examples. After searching on the web I found 1 example in the HBR-2000 transceiver; a project designed and built my respected Canadian colleague Marcus, VE7CA. Click for his web site. I decided to build my own design using a L-match to tune the input to 50 Ω.

The above amp was built around around a 2N2222a. The 39 ohm resistor is not really required with the 2N2222a. For high F-t transistors like the 2N3866, 2N5109 or microwave transistors, ferrite bead(s) or the resistor are *not* an option. Low F-t transistors like the 2N2222a or 2N3904 don't need the UHF oscillation snuffer resistor since they lack real gain at these frequencies. With the design center, you should be able to bias your own amp according to the emitter current you want — choose a BJT, measure or choose its hFE and then choose IE.

Missing from this web page is how to choose an operating point + discussion about DC load lines and related topics. I may tackle these topics on a future web page. I'm not sure anyone cares about this anymore.

The most difficult part was the output transformer. Lacking a base to collector connection, the collector impedance runs quite high and finding a good match into 50 Ω proved impossible — even with a shunt resistor across the primary coil. I saw a strategy in Marcus' amp; AC couple the collector to ground via a 510 Ω resistor. I did this. From then on, it was just trial and error to identify the optimum turns ratio for the collector transformer. An interesting experiment might be to figure out the turns ratio using a lower loss output transformer such as a FT-37-61.

The turns ratio of the various collector and drain transformers on most of these amplifier designs were determined by placing the amp in an output RL measurement setup and adding or removing secondary turns to get the highest possible RL. See the procedures for RL measurement on the RF Workbench pages.



Above — The breadboard of the first version of the base-emitter FBA.

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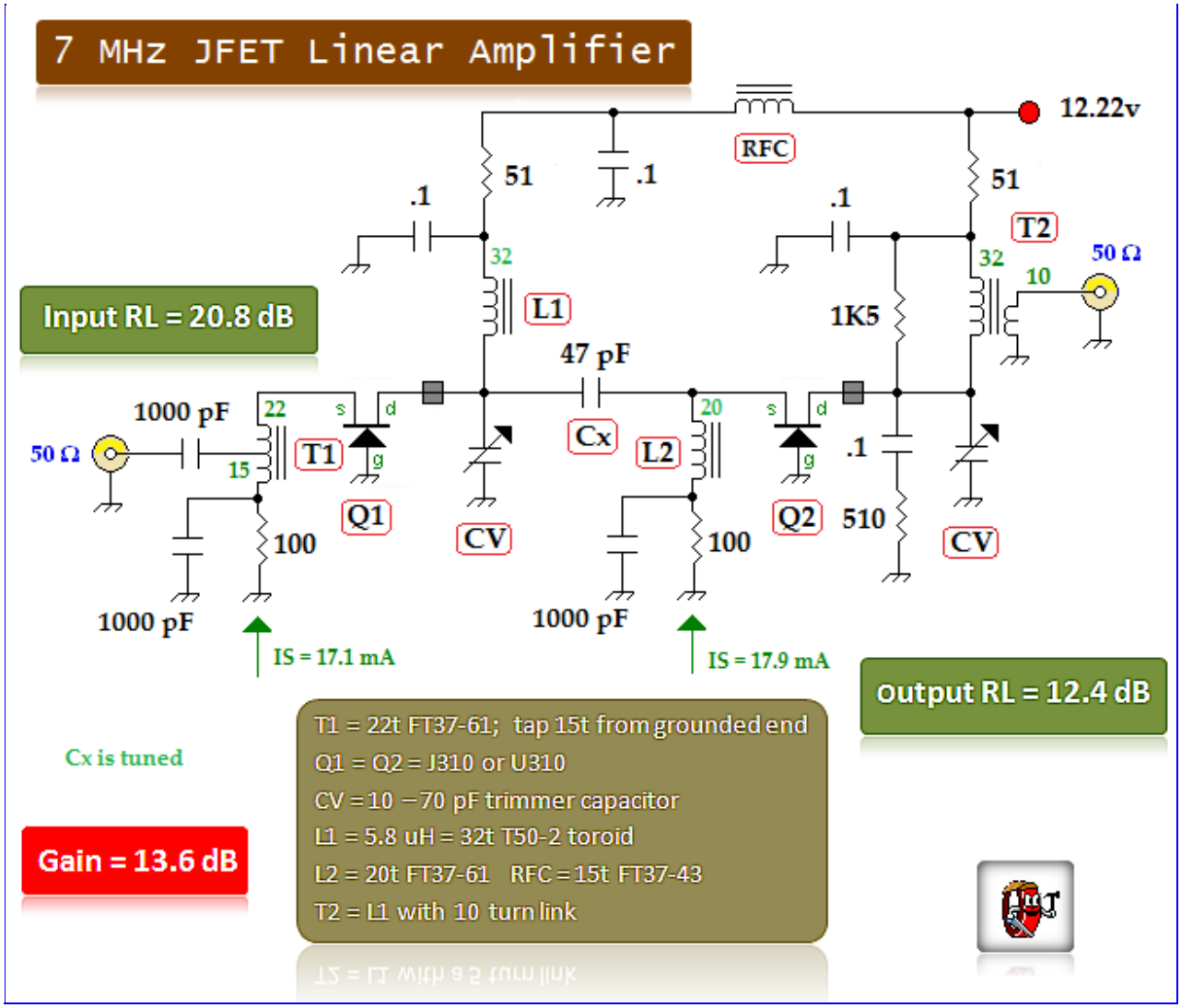
3. JFET Common Gate (CG)Transistor Amplifier Experiments

These experiments focus on setting up a desired input return loss and getting a reasonable output return loss in the CG amplifier. My expectation of an easy set of experiments proved wrong — assumptions never substitute for actually building and measuring.

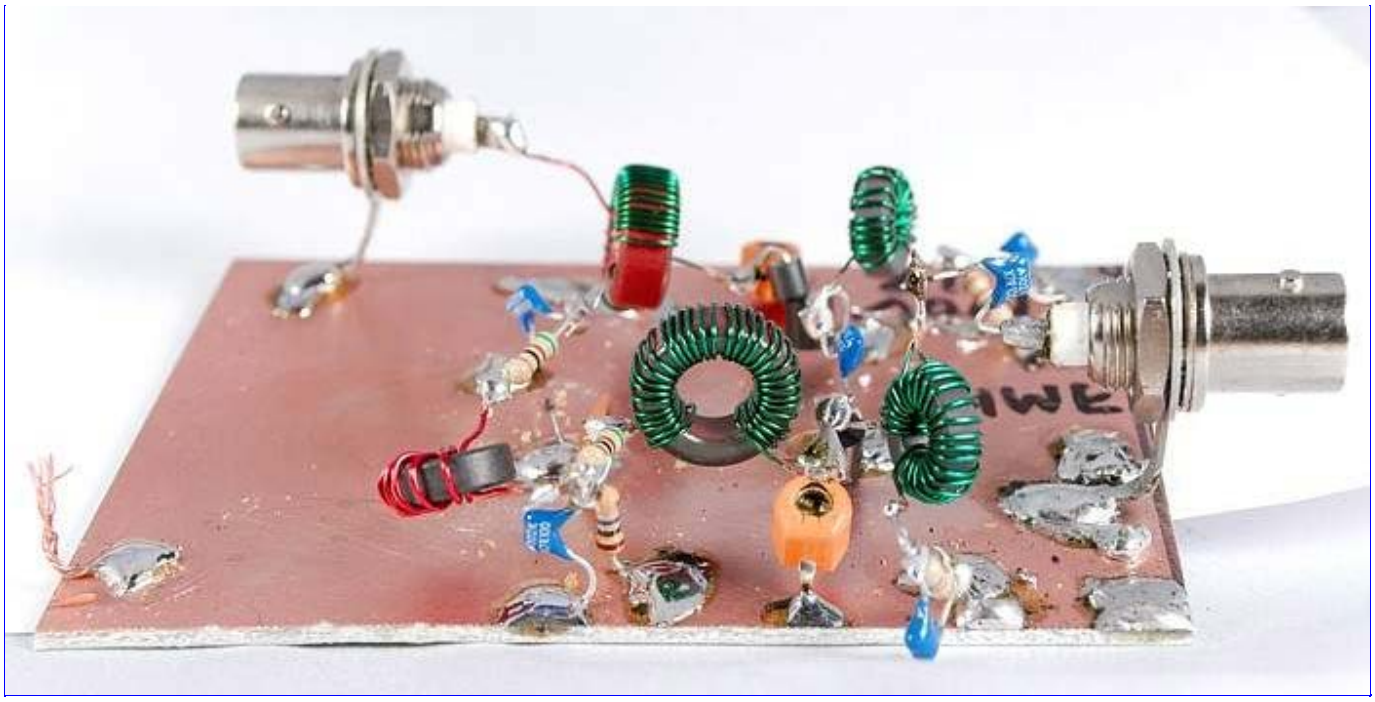
I like motorcycles. The difference between riding a motorbike versus *driving* a car parallels learning on the bench versus learning by just simulating or calculating component "ideal" values on paper or computer. In the car you're isolated from wind, smells, temperature changes and subtle road traction and camber differences that you fully sense on the motorcycle. Bench experiments prove equally visceral and experiential — the sensory input from learning as you build and test circuits imprints deeply in your mind.

1



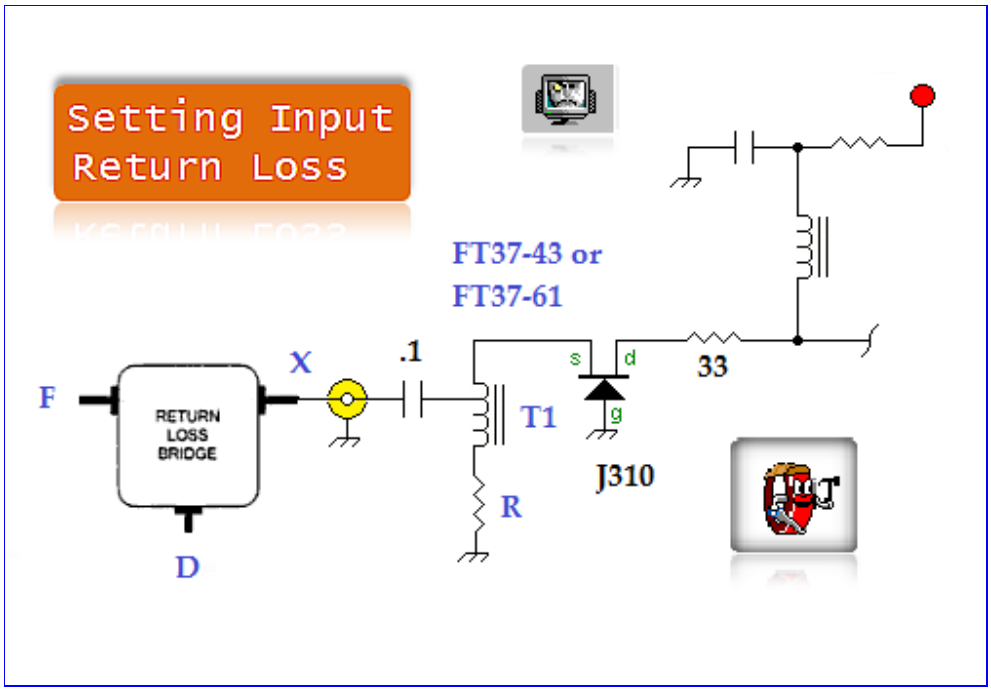
Above — A 7 MHz JFET "linear" amplifier built only for testing ideas — *do not build*. It went through several incarnations and prompted many experiments. The input return loss was deliberately set to 20.8 dB, although I set a RL from 10.0 - 28.6 dB during my experiments. Bypassing the JFET source resistor increases gain, but of course changes input RL.

The output transformer represents a terrible design, but shows the length I went to to try an obtain a decent output return loss. Working with this circuit, led me to abandon tuning the output transformer in situations where a high return loss was desired since the low value resistors required kill the tank Q significantly.



Above — Breadboard of 1 version of the prototype low-level JFET "linear" amplifiers for 7 MHz. Click Click . Cx is tuned with a variable cap and a nearest standard value substituted; in my case 46 pF was the measured value of the variable cap at point Cx.

Setting Input Impedance

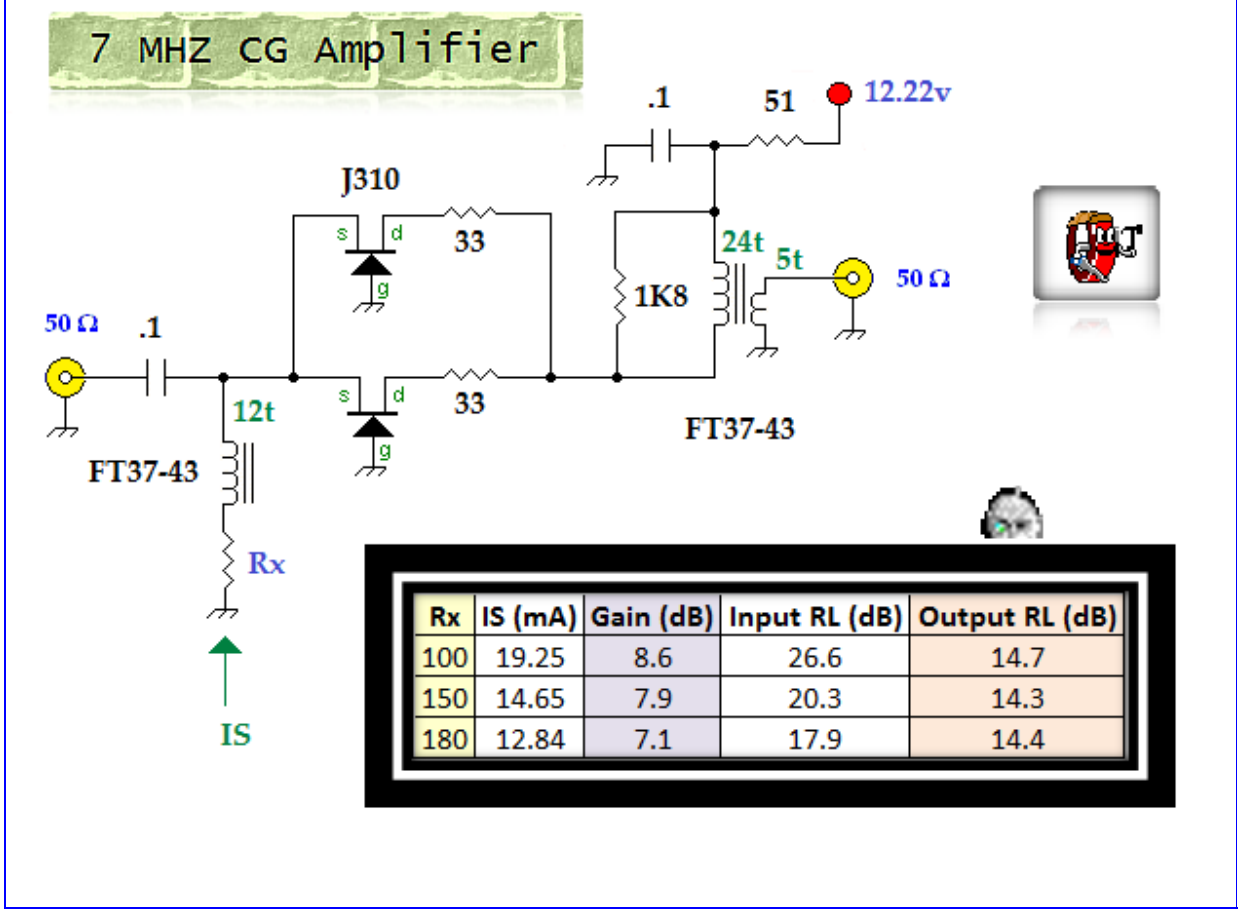


Above — The procedure used to set a desired CG amplifier input Return Loss. Numerous factors influence the input impedance and I discuss them in **#4 Some Factors Affecting Common-Gate Amplifier Input Impedance** on the supplemental web page. I keep some tapped inductors on my workbench such as this FT50-43 or these FT50-61 core inductors. To find the best return loss using such a coil, you can change tap points, remove windings and even wind more turns and solder the 1 end of your new windings to 1 end of the existing wire. Some builders omit inductor taps and manipulate the input return loss other ways as described in the supplemental article.

Normally we set the input match after establishing the output match since the output impedance dramatically affects the input impedance.

Further, you might notice that the tap point may vary between different JFETs. Most of my "real world" coils have at least 2 tap points and I choose the tap that gives the best return loss. More often than not, I bias for 14 - 18 mA and leave off the source bypass capacitor; it's your call.

The input return loss that gives the lowest noise figure is often chosen by engineers.



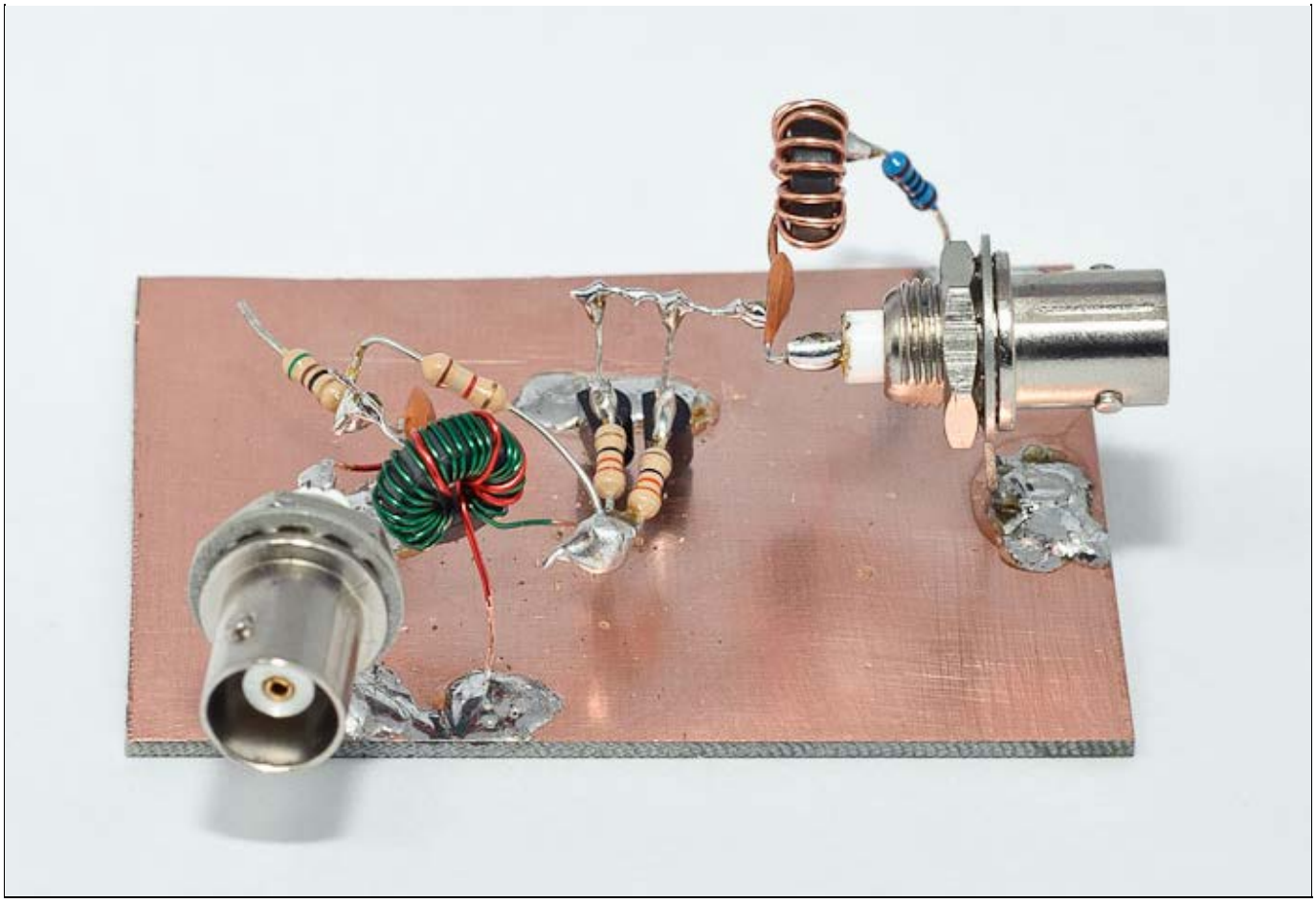
Rx	IS (mA)	Gain (dB)	Input RL (dB)	Output RL (dB)
100	19.25	8.6	26.6	14.7
150	14.65	7.9	20.3	14.3
180	12.84	7.1	17.9	14.4

Above — An experimental 7 MHz common gate amp designed to terminate a diode ring mixer. The best thing about using 2 JFETs is that you don't have to determine the tap point in the decoupling inductor (12 turns on a FT37-43 in this amplifier). I put up to 4 in parallel during my various experiments. It's faster to match just 2 JFETs, so 2 were favored.

The output RL wasn't great at ~ 14 dB, however is probably normal or better than most published amateur projects. I set the output match by adding a shunt 1K8 resistor across the primary winding and then finding the turns ratio to give the best output return loss. Without the resistor, the best output RL will be ~5 dB or worse. The resistor reduces power.

I learned that putting JFETs in parallel in a common gate amplifier reduced the output return loss in circuits using an output transformer like in the schematic above; this is unfortunate.

I wanted an output RL of 20 dB or greater — this is no small request; over a week was spent investigating transformer behavior and finding ways to improve output return loss when you really *want* to.

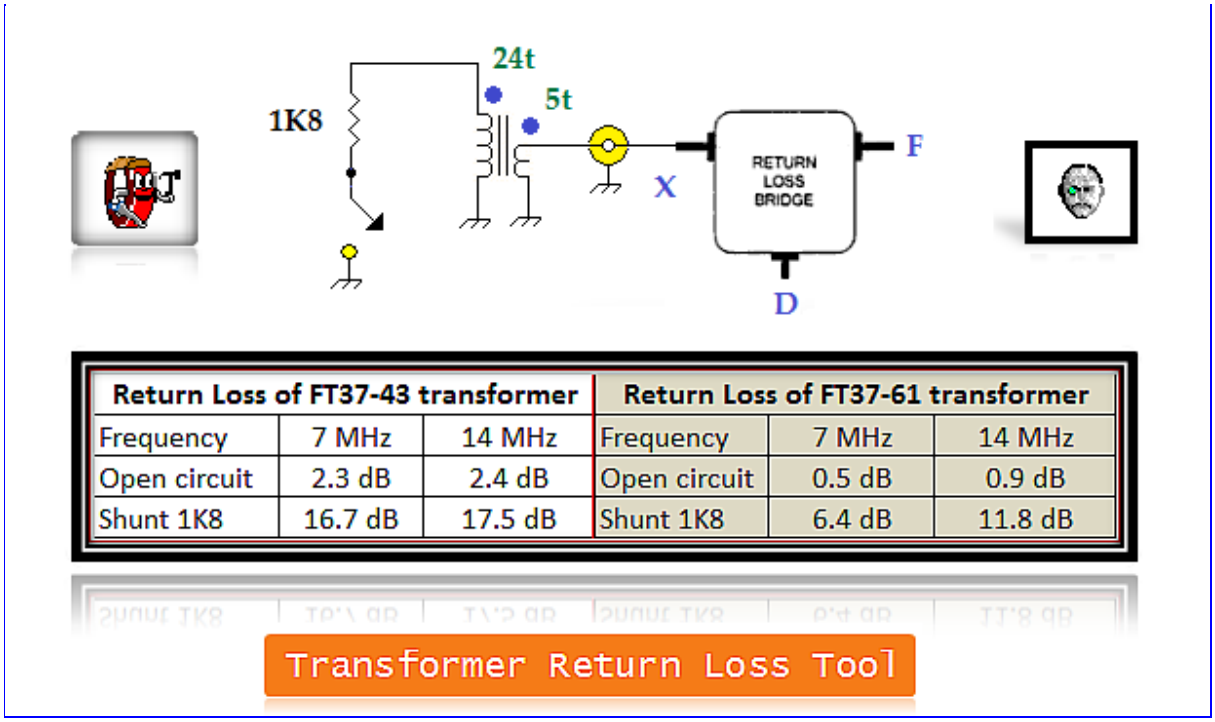


Above — The breadboard of the above 7 MHz CG amplifier.

Output Impedance Experiments

For some reason, I assumed that when using an arithmetically correct turns ratio, the output transformer will end up at 50 Ωs. For example, if I wish to transform 450 ohms to 50 ohms, I'd use a 9:1 impedance ratio (3:1 turns ratio) and get 50 ohms. Sadly, it isn't this simple — impedance transformation is complicated and whole books have been written about it. I'll share some of my experiments that might inform yours.

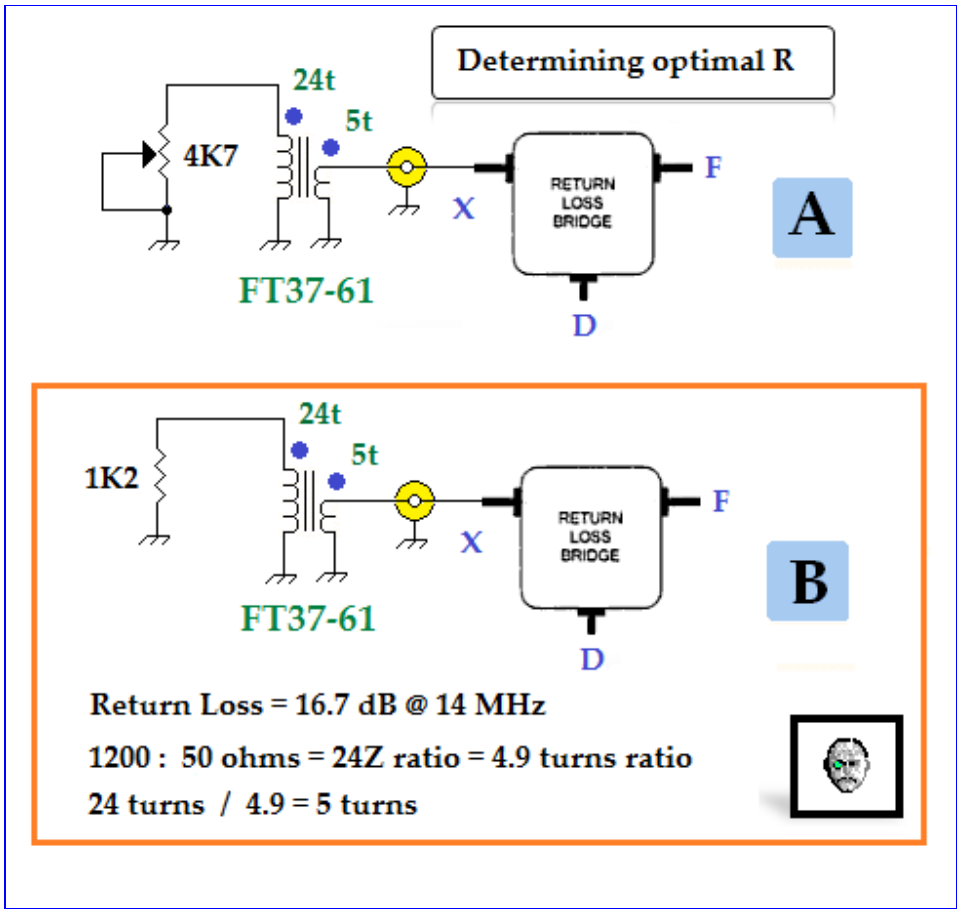
The first task was to built a simple jig to evaluate primary and secondary coupling, turns ratios and return loss.



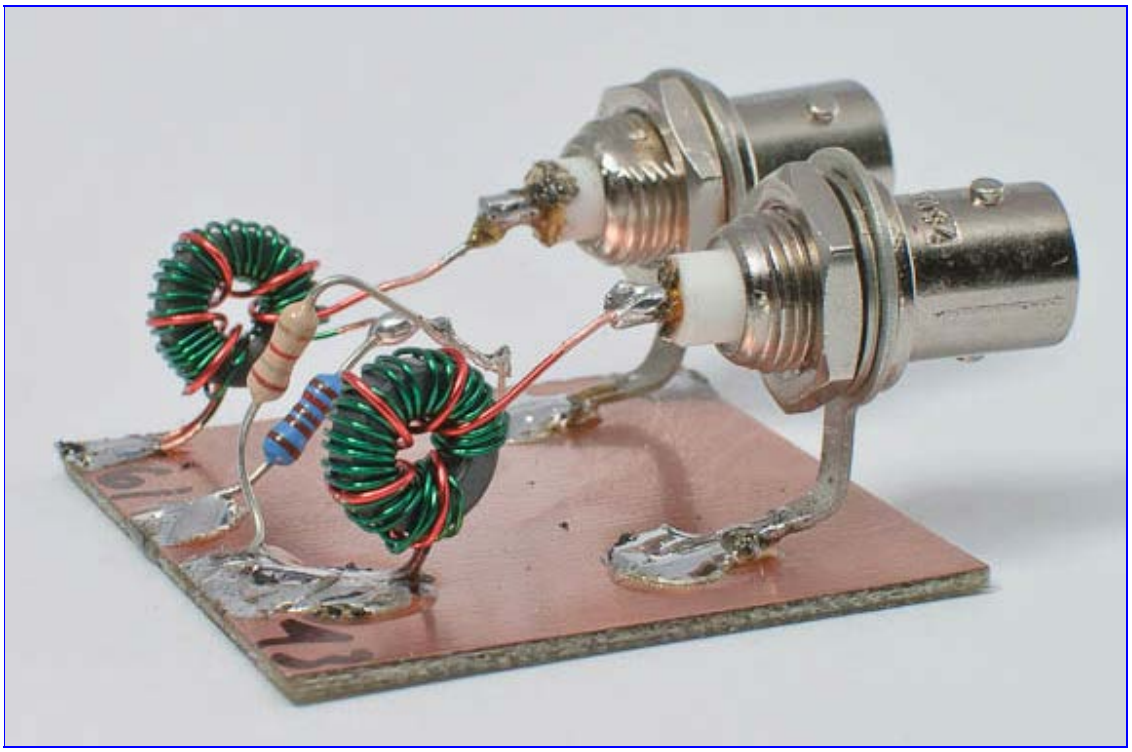
Return Loss of FT37-43 transformer			Return Loss of FT37-61 transformer		
Frequency	7 MHz	14 MHz	Frequency	7 MHz	14 MHz
Open circuit	2.3 dB	2.4 dB	Open circuit	0.5 dB	0.9 dB
Shunt 1K8	16.7 dB	17.5 dB	Shunt 1K8	6.4 dB	11.8 dB

Transformer Return Loss Tool

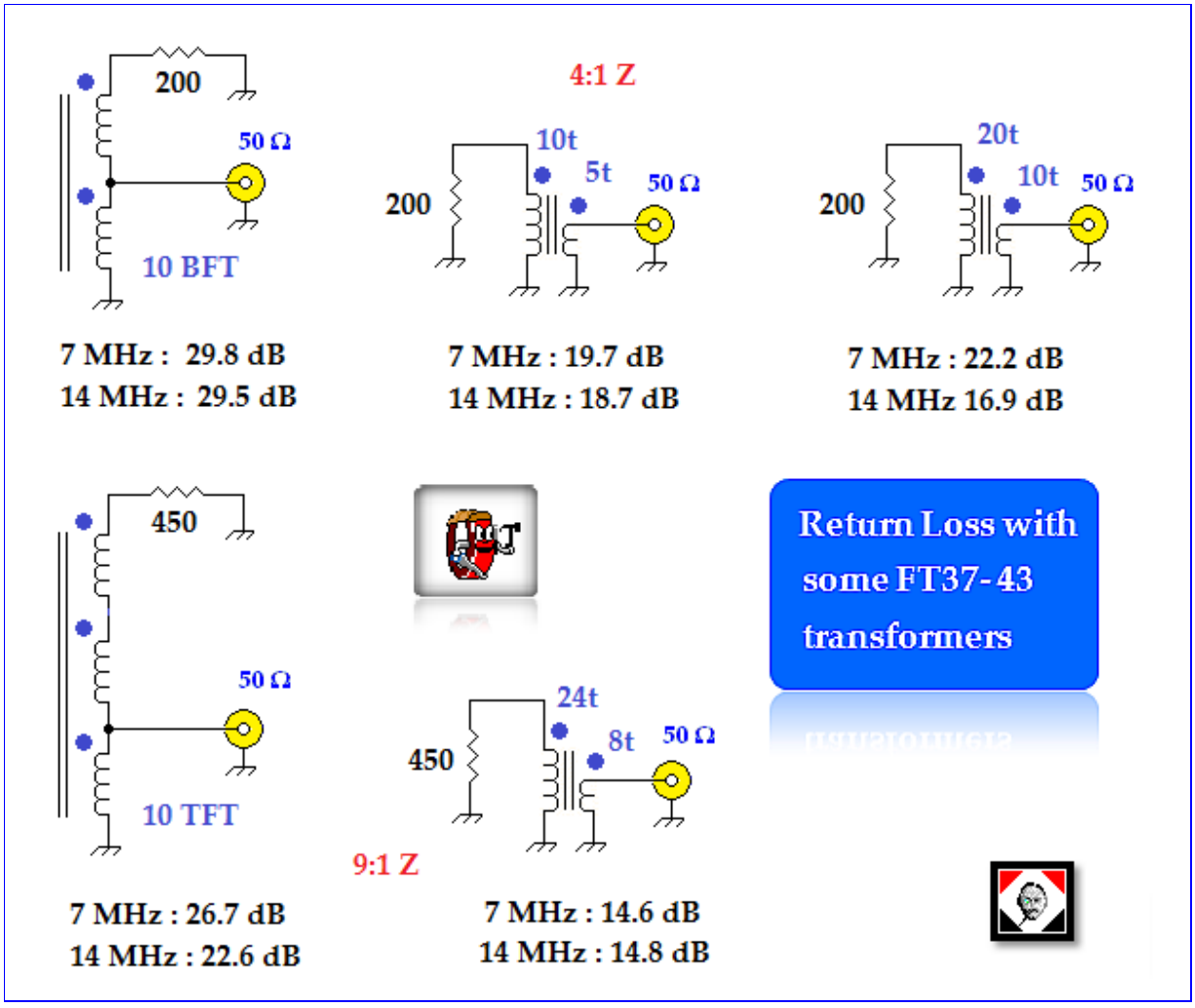
Above — The simple tool built to evaluate the return loss of a transformer out-of-circuit. In this case, I examined the 24t : 5t transformer of the 7 MHz CG amplifier shown earlier. The table shows the best possible return loss when the 1K8 resistor is across the primary coil. Additional experiments were completed and follow below.



Above — An experiment to see if changing the shunt resistor can improve return loss; yes it can. The shunt resistor was a 4K7 potentiometer — Using the potentiometer, I was able to determine the optimal resistance needed to increase the return loss @ 14 MHz of the **FT37-61** ferrite-base 24t : 5t transformer. The pot was removed, measured and replaced with the nearest standard value; a 1K2 resistor. The best possible RL was 16.7 dB using a 1K2 shunt resistor. At 7 MHz, the FT37-61 didn't work well. Five turns on a FT37-61 based transformer doesn't have enough inductive reactance to get a good return loss.



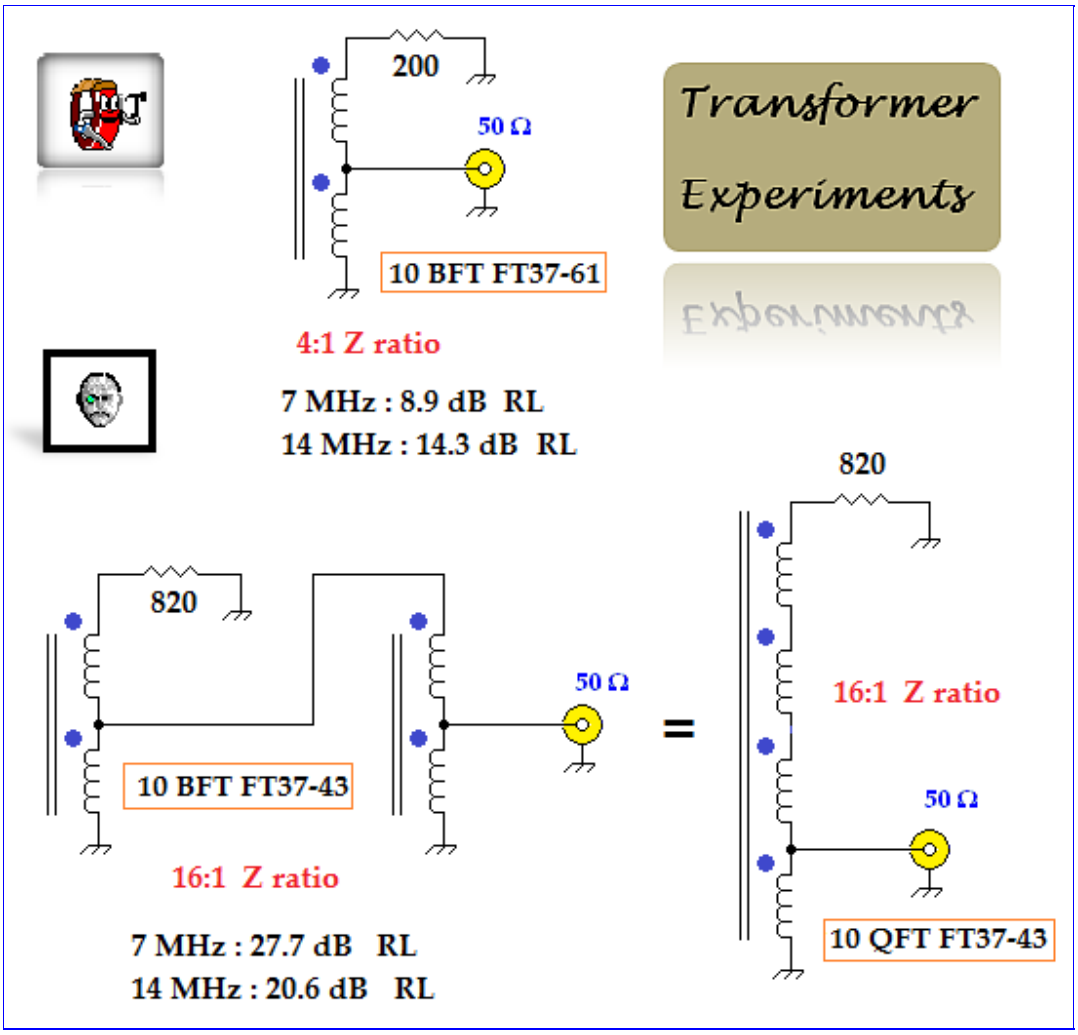
Above — The transformer testing jig. I omitted the switch shown in the schematic above and just soldered the shunt resistor across the primary winding.



Above — Some of the outcomes using the transformer jig pictured above. While I basically understood that transformer efficiency tends to fall as

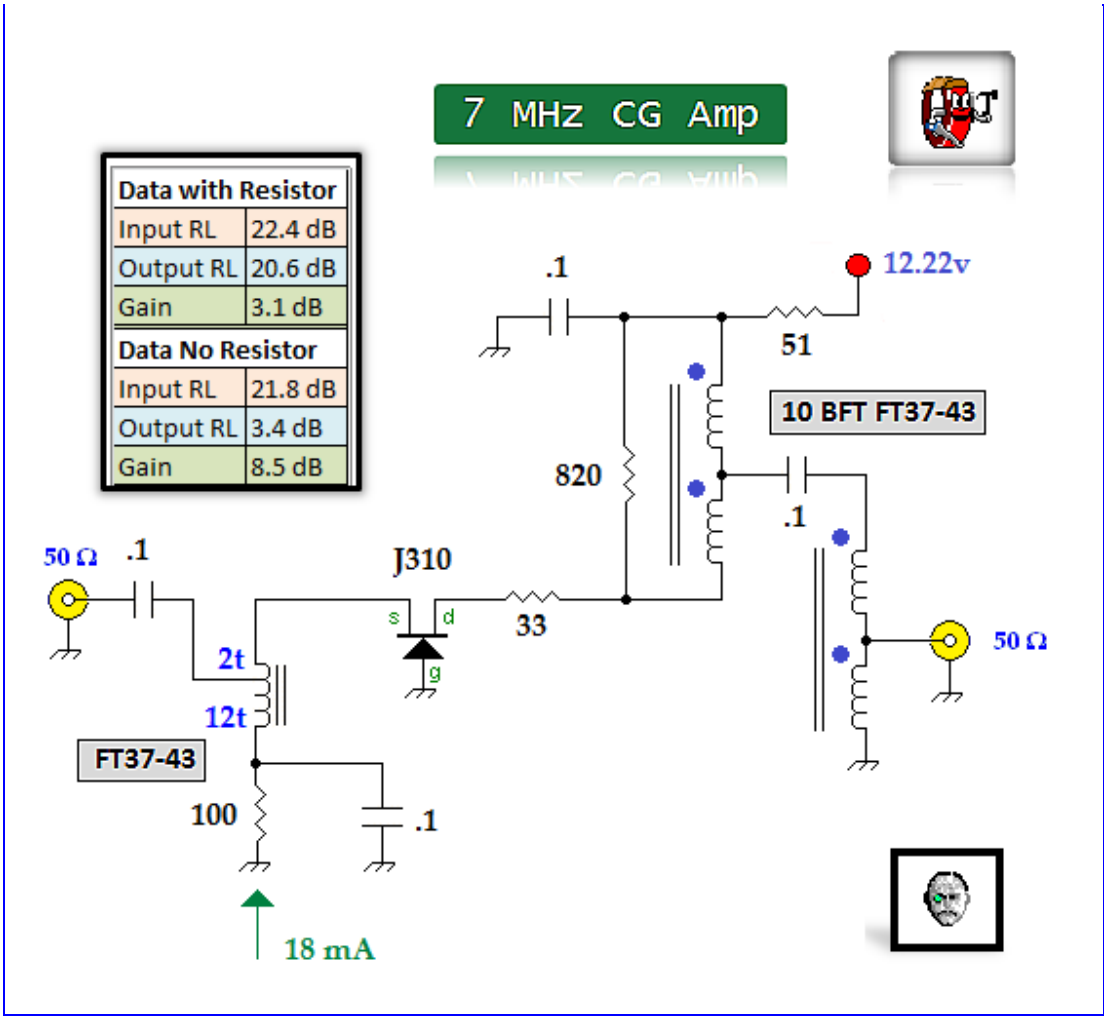
the turns ratio increases, I never thought this would also happen with return loss. By *no* means do these crude experiments constitute science, but the following themes emerged:

1. The better coupling of transmission line transformers (bifilar, trifilar etc.) translates into improved RL over conventionally wound transformers
2. Limiting the turns ratio to 3:1 or less generally improved the return loss. As the turns ratio moves above 3:1, the best possible return loss tends to decrease.
3. The smaller or secondary winding should have 4-10 times the inductive reactance of the impedance it's connected to. For a 50 ohms impedance this means a minimal XL of 200 - 500 ohms. I noticed a weak trend towards better return loss with higher XLs. This means that to use a FT37-61 at 7 MHz, the secondary winding should be 9 -14 turns or so.



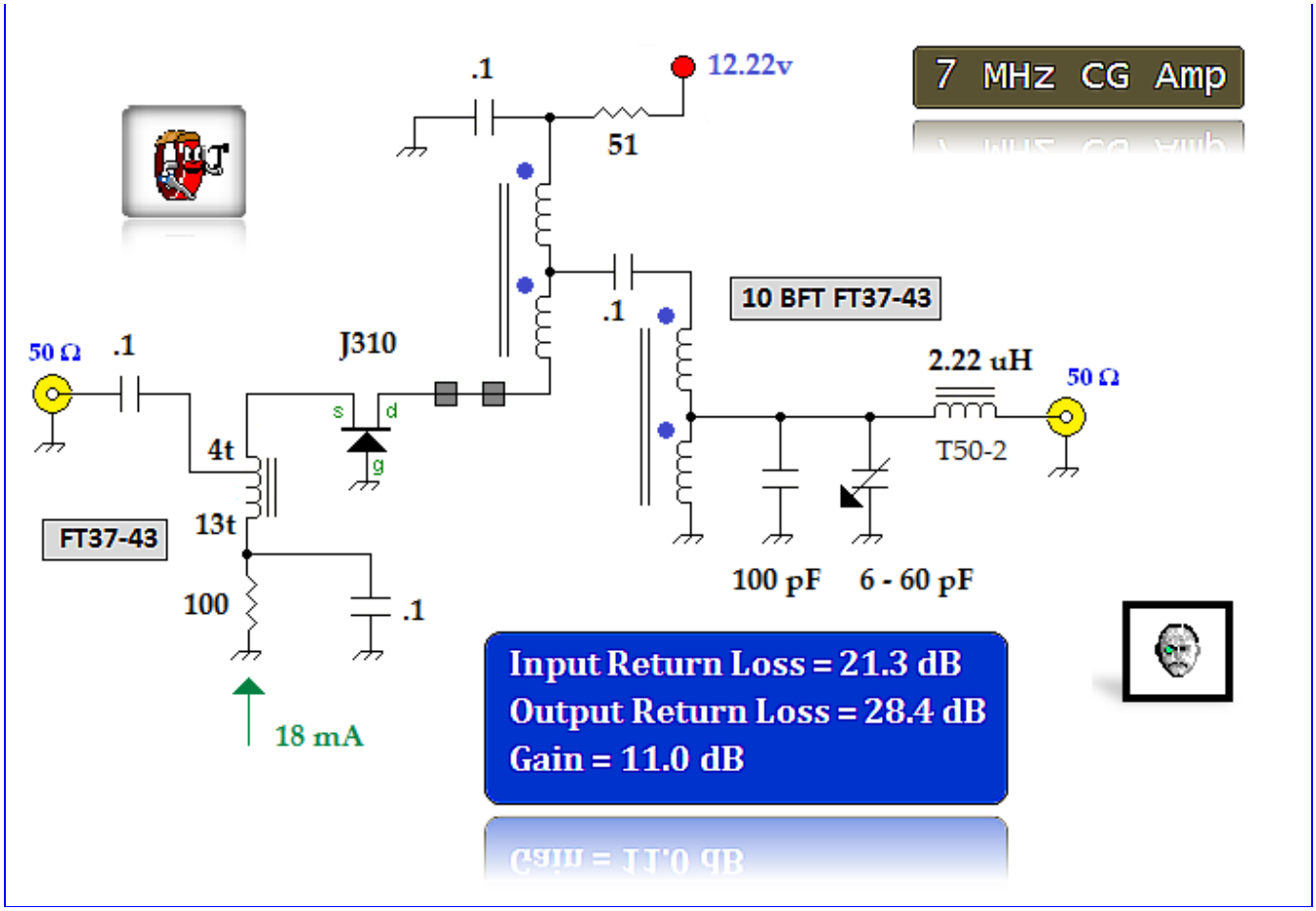
Above — Further transformer experiments. For a 4:1 impedance transformation at 7 and 14 MHz, a FT37-43 ferrite toroid gave a better out-of-circuit RL than the FT37-61 *The comparison transformer with a FT37-43 ferrite core was shown earlier*. It's possible to transform a big impedance such as 16:1 by cascading 2 bifilar transformers, or by using a quadrifilar transformer. I didn't build the quadrifilar transmission line transformer, but show it for completeness sake.

Of course, once you connect the transformers to a real circuit, things will change — still it's great to be able to examine transformer return loss in a controlled environment.



Above — a common gate amplifier experiment using 2 cascaded 4:1 Z transmission line transformers. Data with and without the 820 Ω resistor shows that while the resistor gives a great output RL; it eats a lot of power. In cases where I've seen cascaded transmission line transformers used, the resistor was omitted. The Ugly Weekender transmitter by Wes, W7ZOI provides a good example.

In many cases, it's prudent to sacrifice gain for return loss, however, when you see a builder (like the former me), put a 32:3 turns ratio on a 5 MHz amplifier output transformer and label the secondary windings **"50 Ω",** we'll know better.



Above — An evolution of the amplifier above to get the best possible output RL. I omitted the 820 ohm resistor and matched the output with an L-network. The return loss on the output of the second transmission line transformer (measured before the L-match was added) was 3.4 dB.

The L-match values were *roughly* determined by using this chart (you can also do the math). According to the chart an *(output)* RL of 3.4 dB, is either 10 ohms or ~250 ohms or so. Ten ohms is unlikely, so I designed my L-match to match 250 to 50 ohms. This provided some starting values for the L and C parts and the rest was done on the bench using trial and error with an RL bridge. At the time, this was the highest output RL I'd ever achieved.

RF engineers use math to calculate impedance (they always do). I sent the schematic to Wes, W7ZOI for his analysis and summarize his return email comments as follows: At 7 MHz, the XL of the 2.22uH inductor is 97.6 Ω, therefore the impedance looking into that with 50 Ohms as the load is 50+j97.6. A complex inversion of this value gives a complex admittance that has a real part: 0.0041. Flipping that gives 240 Ω. The equivalent reactance is inductive with a value that would be tuned by a 184 pF capacitor; a bit more than you have there — so there is some reactance presented by the center tap of the second transformer. Neglecting these details, the L net generates about 240 Ohms. The two transformers then kick the Z up by 16 to 3856 Ohms.

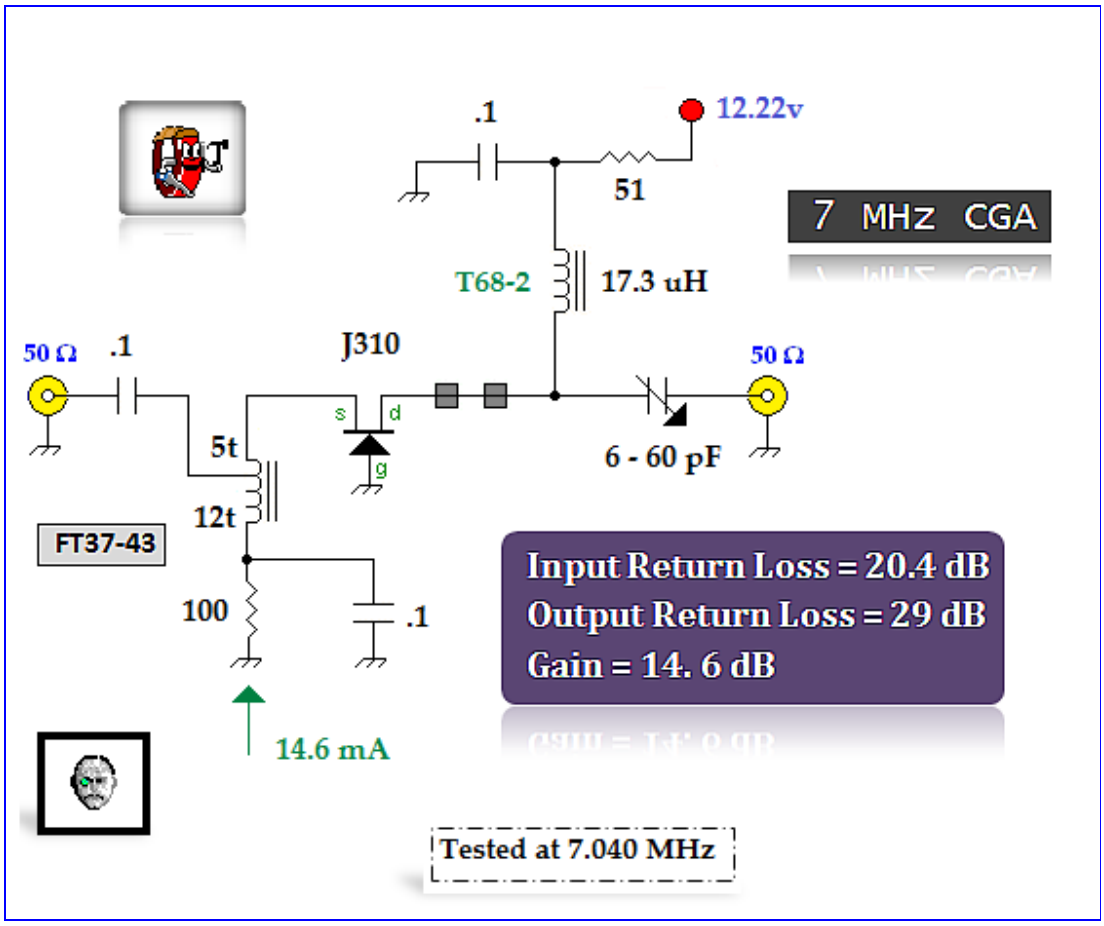
I was pleased that my simple chart gave a value close to his calculation. Test it out — the chart may work okay for you.

The input match is over 20 dB and reasonable. More time could have been spent on the input autotransformer by tapping and such to increase the input RL, however, time is the 1 resource we all seem to lack.

The final amplifier experiments employ an L-match to set output return loss. When reading electrical engineering books you'll often see all sorts of matching networks on both the input and output of FETs and BJT amplifiers. The networks look simple, but in practice, aren't. They tune sharply, have a low bandwidth and in the case of the CG amp, harbor a big problem — tuning the output for the best output return loss, **dramatically** affects the *input* return loss and potentially, your return loss measurement by the reactance affecting the RF signal in your bridge.

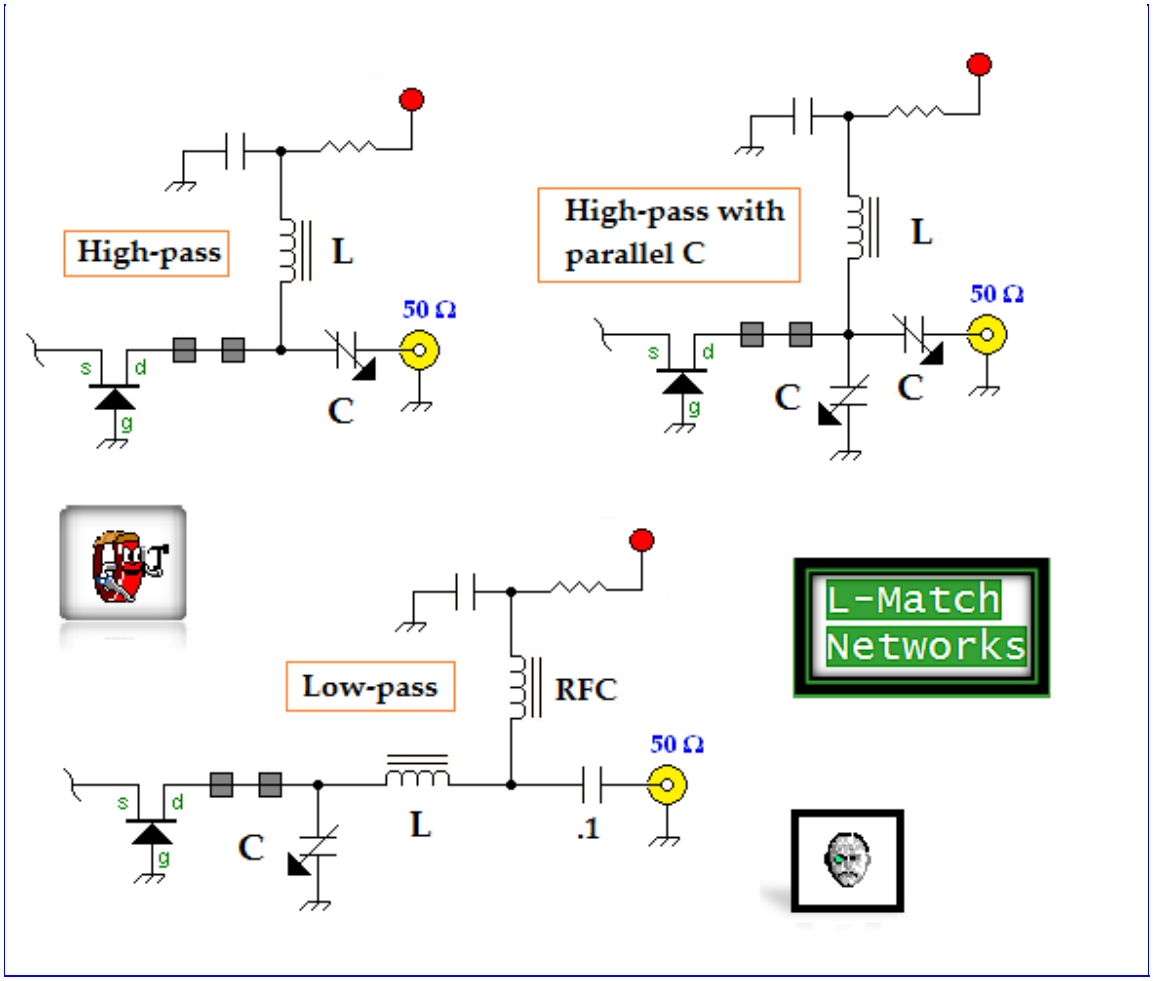
When tuning the output, you're actually changing 2 complex impedances — this is not trivial.

Also if you're off by a few pF or tens of uH in your network C and L values respectively, you can wreak havoc with the measurements. At this point, I don't possess all the skills needed to tune both the output and the input network to a RL of 20 dB or greater; especially with a broadband input.

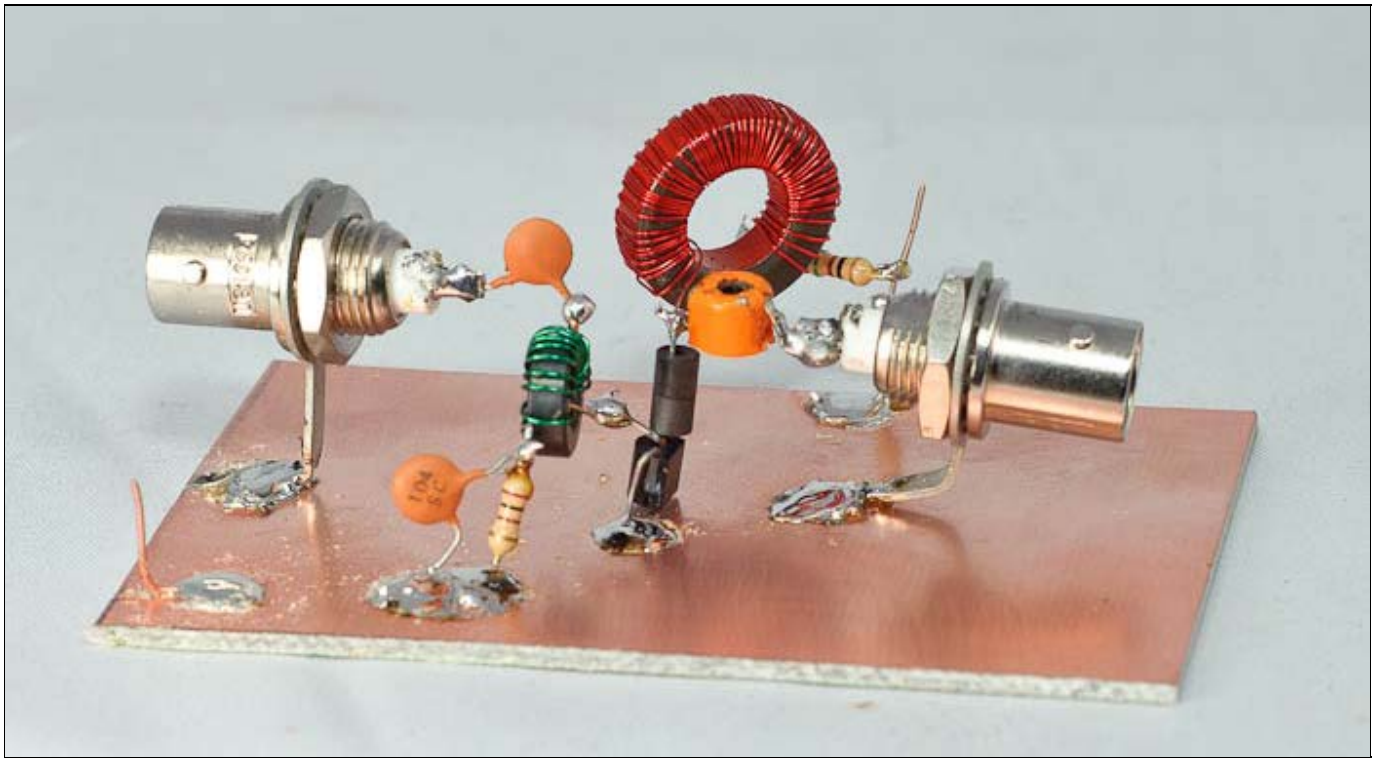


Above — A common gate amp employing a high-pass L-network to match the output. Miraculously after 2 hours of tweaking, I obtained a good input and output match; however this amp isn't reproducible. The inductor was wound on a T68-2 using 28 gauge wire — always a pain. Through trial and error, I learned that the output impedance of the drain was around 11800 ohms. Starting with 18.4 uH on theT68-2, I removed 2-3 turns at a time until a reasonably low return loss was obtained; then I removed 1 turn at a time. I went too far and had to add back a turn. I clipped the excess lead every second turn which made it tedious, but exacting. It seems that the L value is very critical – it would be nice to use a variable inductor to figure these things out. Compressing and expanding the windings also provided a simple way to vary inductance.

In several other circuits, the best possible input return loss was only 14 dB. Mistuning also caused oscillations to occur in one 14 MHz amp with an output network inductor of 7.4 uH. I also tried a 14 MHz amp with an L-match on both the input and output, however, was unable to match both the input and output due to the interplay between them.

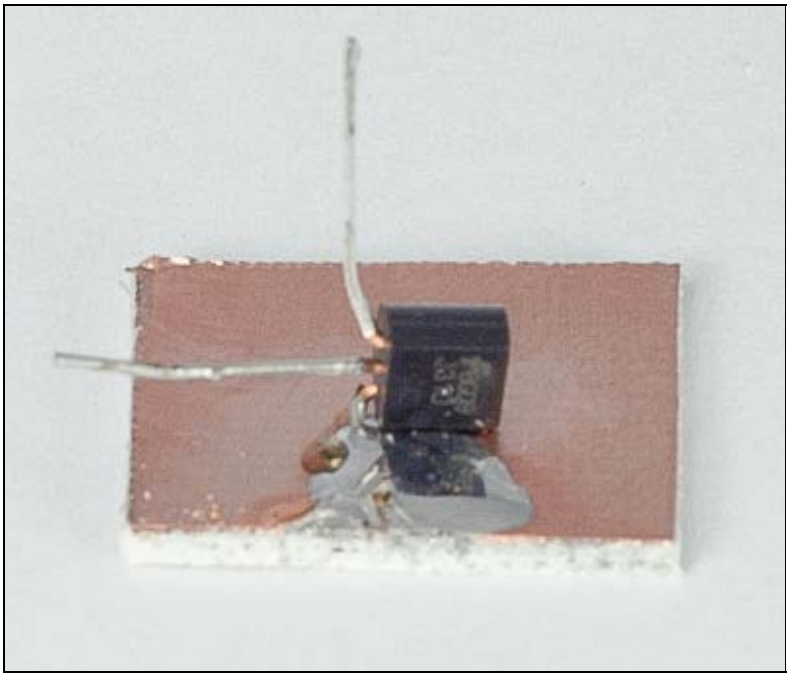


Above — Here are 3 possible L-network configurations for tuning a CG amplifier output. They can be used in other circuits and are worth studying. The L-match with 2 variable capacitors generally requires lower inductance than the others.



Above — A breadboard of 1 of the high-pass tuned CG amps. The gate lead on this transistor is too long — the inductance will likely cause UHF

oscillations. 2 ferrite beads were placed on the drain to mitigate these, but a better construction technique is recommended and shown below.



Above — the preferred way to ground the gate with the JFET on its side. The hole-through version of the U310 JFET has a metal case that is connected to the gate that makes it ideal for grounded gate amplifiers. Some suppliers only sell SMT versions of the U310 now.

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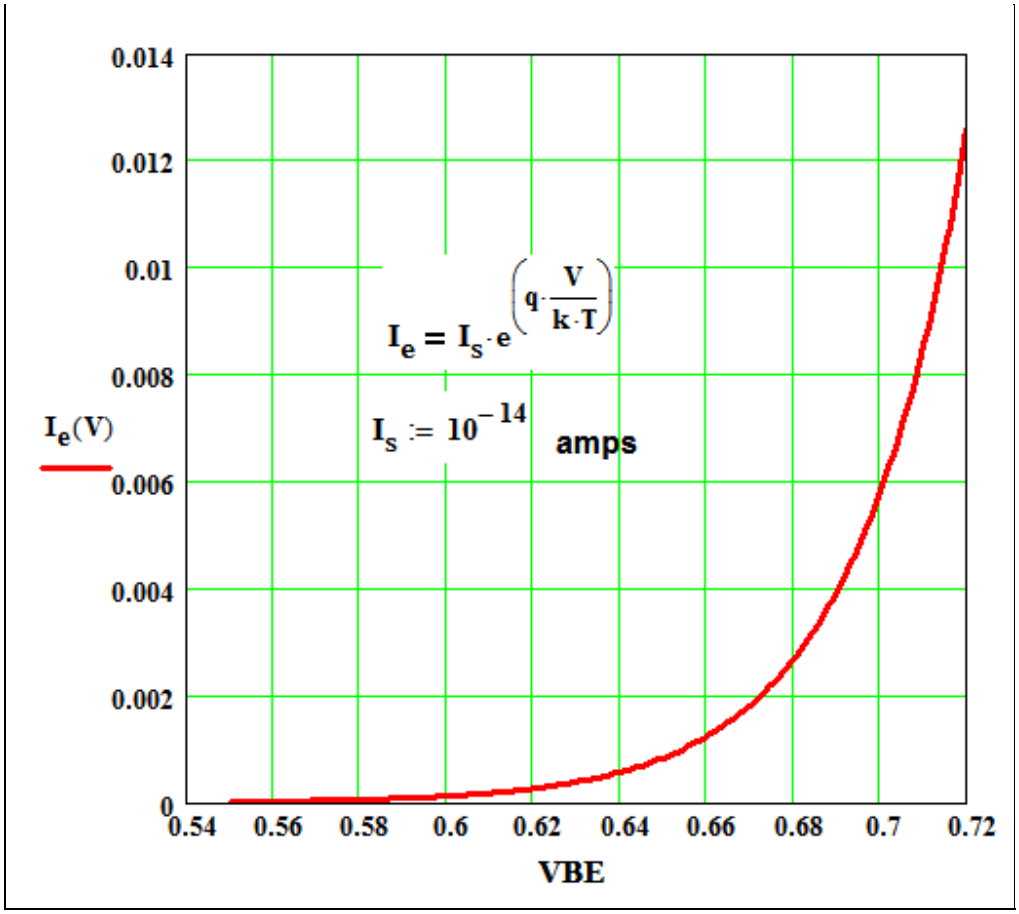
4. QRP — Posdata for January 2013: Transistor Bias Model

This discussion concerns setting up the DC bias point for linear BJT operation.

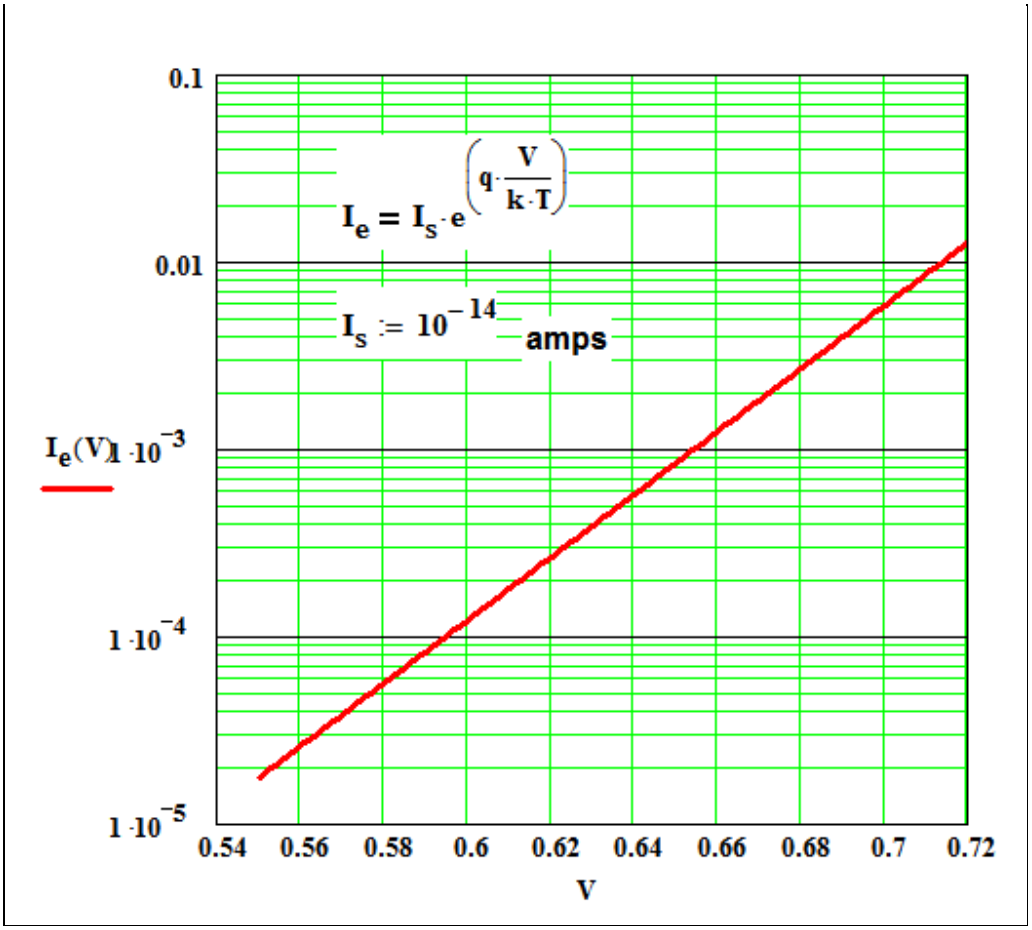
Earlier I stated that a bipolar transistor is a voltage controlled device. A few readers thought I made a typo: something I frequently do, but not in this case, since I purposely made that statement. In reality, the argument could go either way since collector–emitter current is controlled by the base-emitter current (~a current controlled device) and by the base–emitter voltage (~a voltage controlled device).

Stated using the correct physical model, a transistor is a **current controlled current source**. With external circuitry we can manipulate this physical model into a voltage controlled current source, or a voltage controlled voltage source, or even a current controlled voltage source. Whether you model the transistor with current or voltage, the math tells the truth when properly examined.

Please view the following two 2N3904 SPICE models generated by Wes, W7ZOI for me many years ago when I began to learn small signal analysis using impedance and hybrid parameters, plus set out to learn ways to establish DC bias and temperature stability in BJTs.



Above — the Y axis shows how changing base-emitter voltage or current changes the VBE. We tend to assume a VBE of 0.7, however, the math shows the truth. Whether we plot voltage or current for the Y axis data, the graph slope remains similar. The greater the applied DC voltage placed on the base-emitter port, the more current will flow.



Above — Logarithimic base current plotted against VBE. If we want this current to increase, we need to put more DC voltage on the BE junction. On the bench, we may easily measure base voltage to confirm our calculations — measuring base current proves more difficult. Whether I'm setting up amplifer bias with voltage dividers, a current source, or even biasing it with a downstream AGC voltage, I prefer to think in terms of voltage control — although I get that V and I truly just coexist.

Current or voltage modelling — it's your choice and the math will guide you. Look for these equations on the Web, or in second hand bookstores. I've got 8 or 9 transistor theory books now and they're really timeless.

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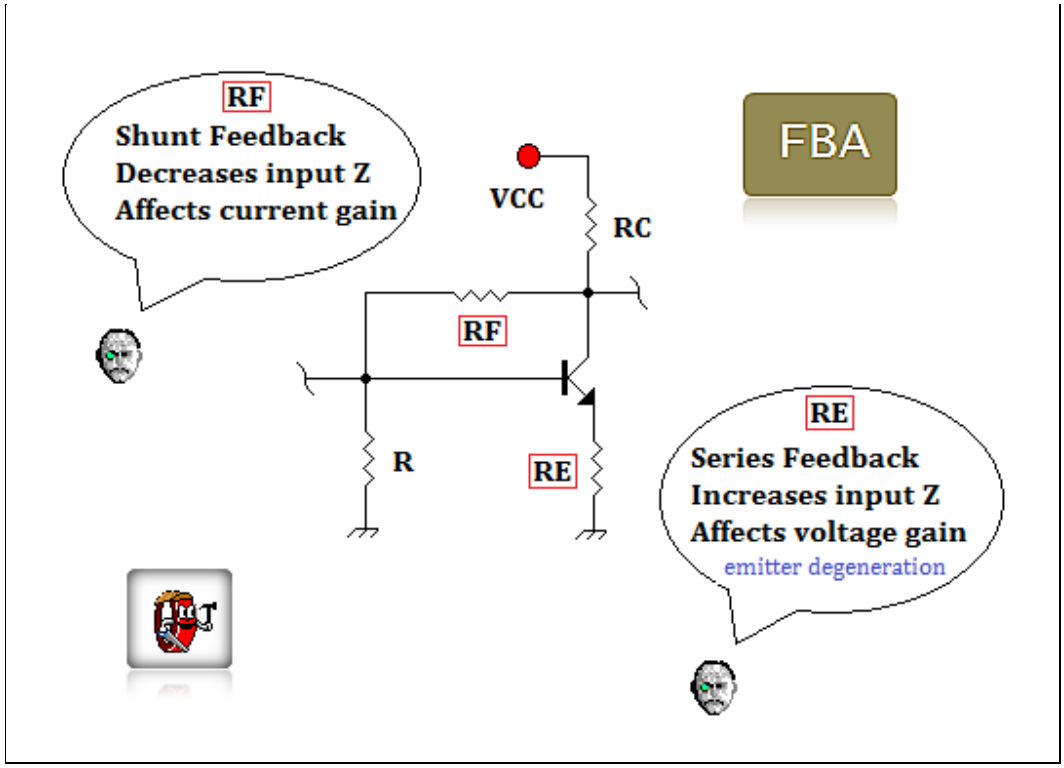
5. Miscellaneous Figures and Photographs


RF
Shunt Feedback
Decreases input Z
Affects current gain
FBA
VCC
RC
RF
R
RE
RE
Series Feedback
Increases input Z
Affects voltage gain
emitter degeneration

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6. References

1. Communications Receivers - Principles & Design. Rohde, L.R and Bucher, T.T.N. 1988. McGraw-Hill.
2. Technical Correspondence QST Magazine (ARRL) Aug 1990. Hayward W.
3. Experimental Methods in RF Design (ARRL) 2003. Hayward W. Campbell R. Larkin B.
4. Microwave Handbook Vol 1. Components and Operating Techniques. 1989 RSGB.
5. Emails with Wes, W7ZOI and Professor Ken Kuhn, Winter 2010-2011 — Thanks! Никогда не забуд.

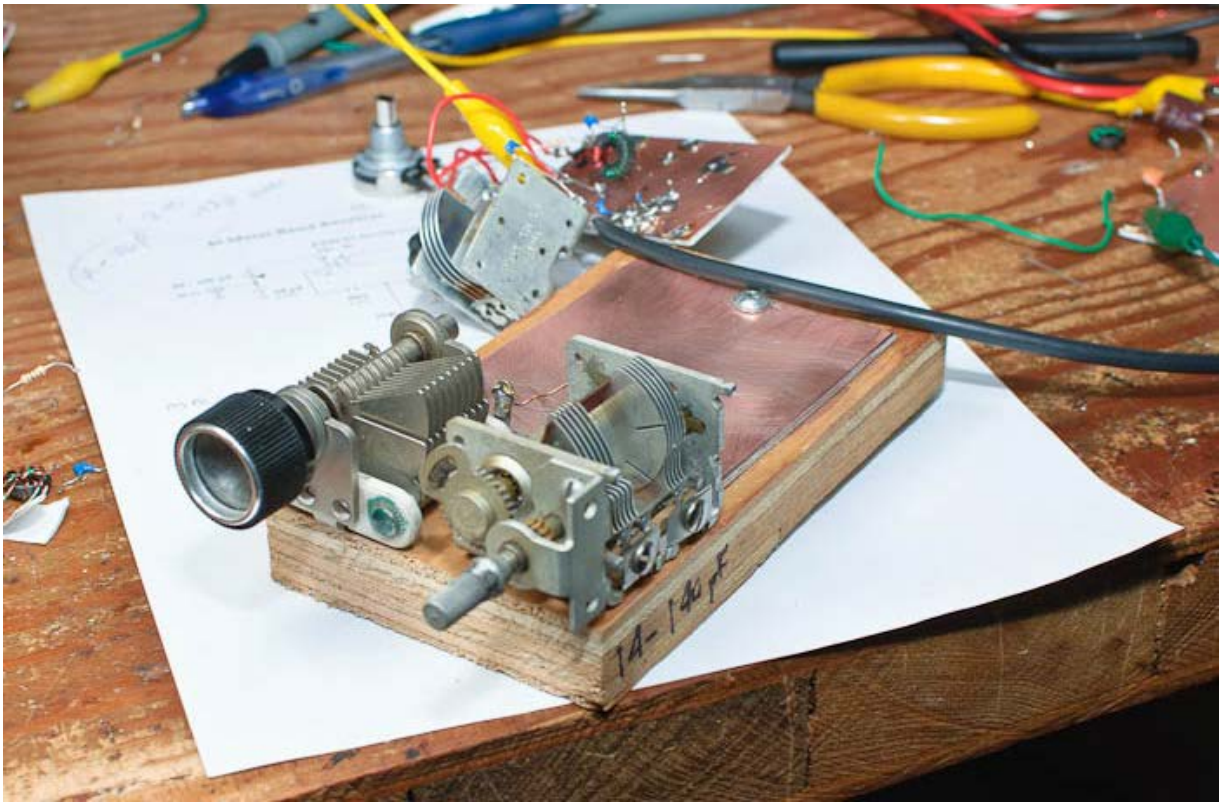
Hobby and Fun 2011

Introduction

Hobbies are supposed to be for recreation. Electronics should be fun, not stressful — heavy math, big parts counts and complexity are more likely to scare away experimenters than recruit them. This page avoids the measurement focus of my latest stuff and simply promotes fun and discovery.

You might be interested to know that my simple experiments/projects garner the most emails. Many wrote "I'm rediscovering electronics", or, "I want a simple and fun hobby". Hobby and fun are my goals too.

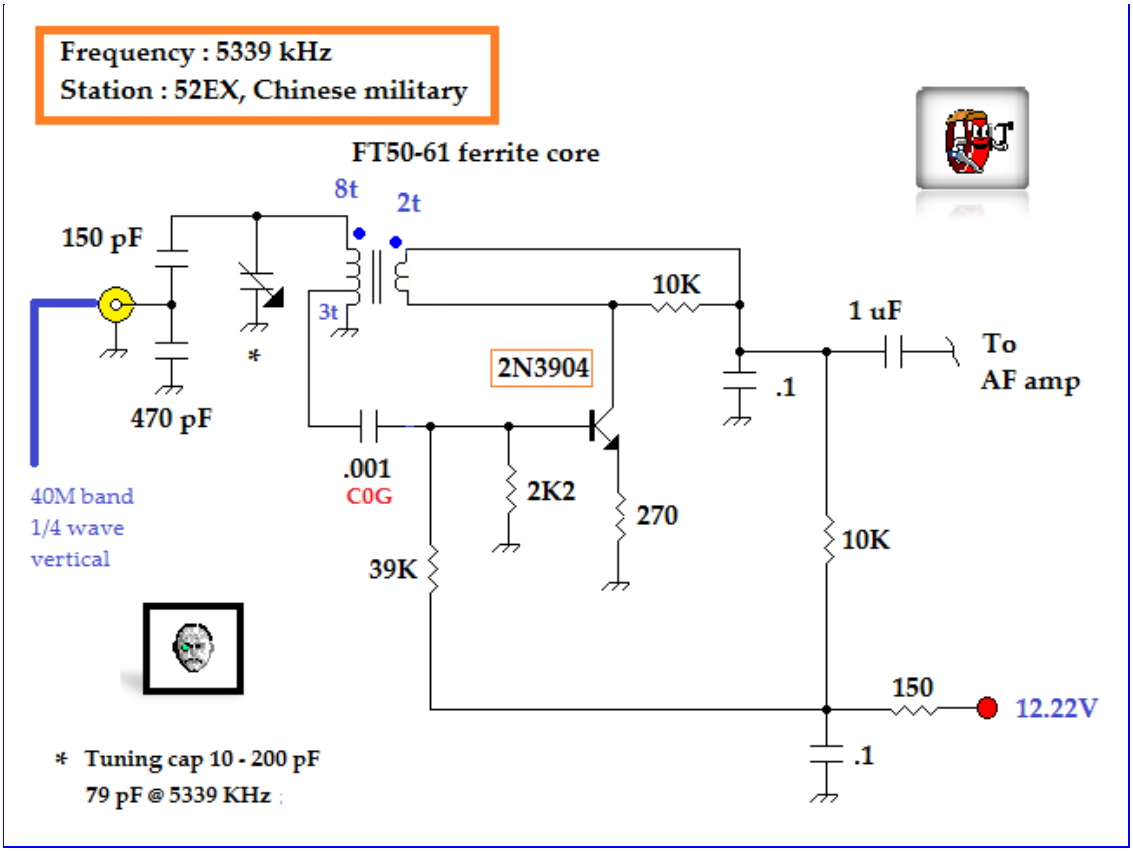
Simple Regenerative Receiver Experiments



Above — 2 air variable capacitors and a copper clad board screwed onto a piece of wood for my bench musing. Regenerative receivers delight and amaze — some builders take them very seriously. I respect this, but to me; they should be as simple as possible. I wanted a 2 stage "genny" receiver for this page and present 2 different receivers; 1 is my design, the others is a JFET variant of a favorite W7ZOI circuit. Quoting Wes, W7ZOI "feedback your imagination".

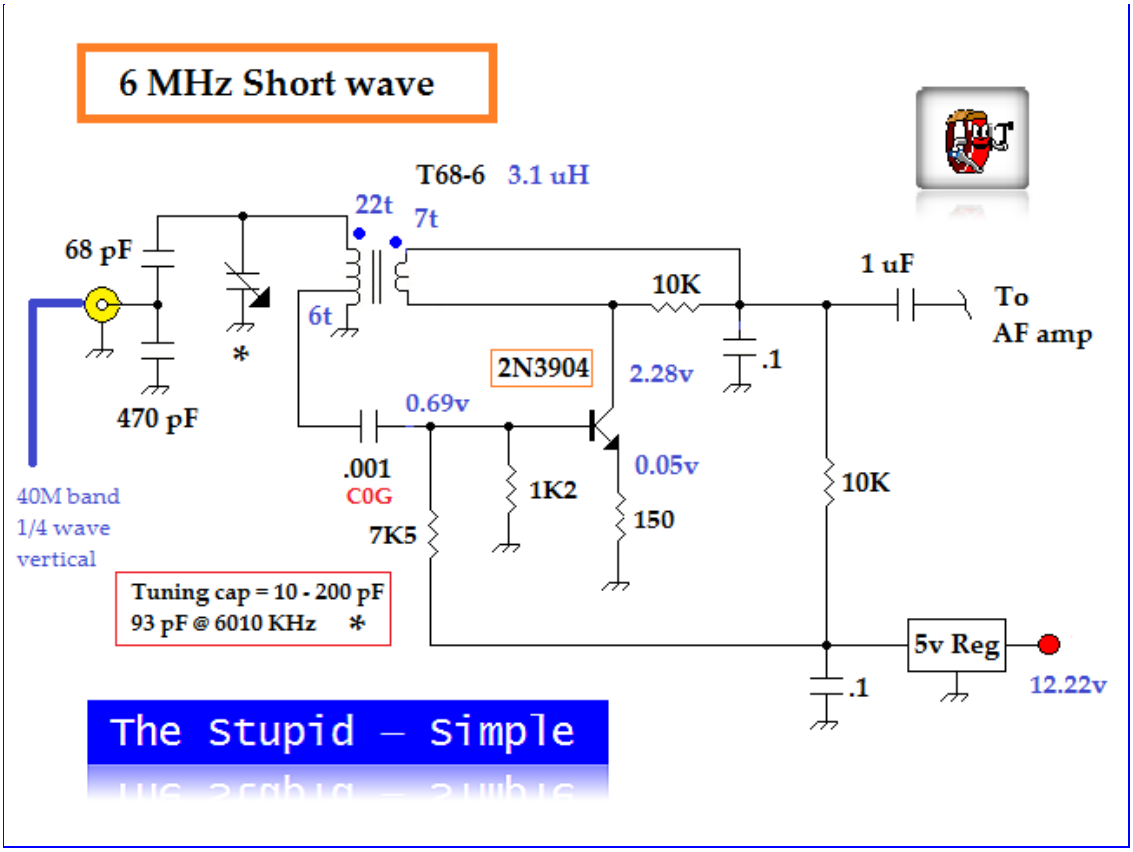
Some builders place a simple common gate or a common base RF preamp on the input to boost gain and reduce antenna radiation of the RF oscillator, while others place an RF gain control on the input — usually a potentiometer; to prevent overloading the RF stage. I won't prescribe what to do — that's up to you. There are countless example of regenerative receivers on the web and you can many spend hours viewing them. Some of the most intriguing are those built by Russian speaking experimenters. Example link.

My circuit ideas are meant as fodder for your own experiments.



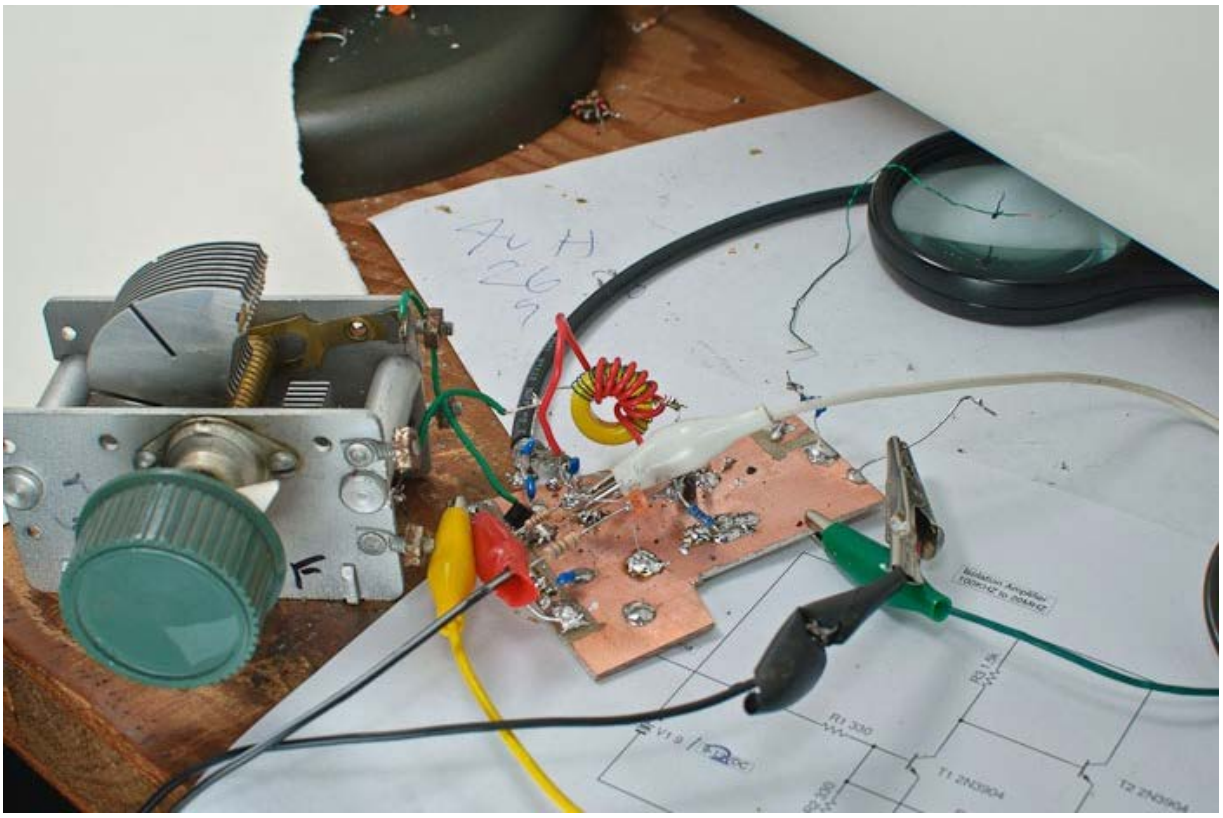
Above — An experimental, ultra-simple "CW" receiver. At 5339 KHz I hear strong Morse code each night. It's suggested to be from China, but I'm unsure. Connected to my 1/4 wave 40M band vertical - a simple matching network and trimmer tuning capacitor were fitted to the input. Here's some audio. I like the beat note of this receiver - it has no regeneration control and is fixed for CW. Minimalist circuits are fun — some hardcore regen builders might freak out; no voltage regulator (here's a version with that + a T68-6 inductor), no regeneration control (here's a version with that), a relatively low Q coil etc. I wanted to try my hand at design and not just copy someone else's receiver.

I call it the Stupid— Simple receiver and although it emits crisp, warm audio, some bench work is required to get the correct bias and appropriate amount of positive RF feedback — an experimenter's circuit that explores DC bias and AC feedback. The sort of thing a father can build with his son. We need more circuits fitting this profile.



Above — The Stupid—Simple experimental receiver set up for ~6 - 7 MHz. 2011 marks the 50th anniversary of Radio Habana Cuba. We tune RHC at 6010 KHz, and as long as I've been listening to SW radio, I've tuned this station. Here are 3 sound bytes from around 6 MHz recorded from 2:30-2:40 GMT on Feb 28, 2011, including an old repeated episode of DX'ers Unlimited by Arnie Coro which had faded out by the time I located and turned on my audio recorder. Audio1 Audio2 Audio 3. The audio stage in these recording was a discrete transistor AF amp I designed, however, an LM386 was chosen for the final amp to keep the parts count and difficulty down. I got a little too close to the receiver a couple of times during the recording and made the open circuit breadboard squeal.

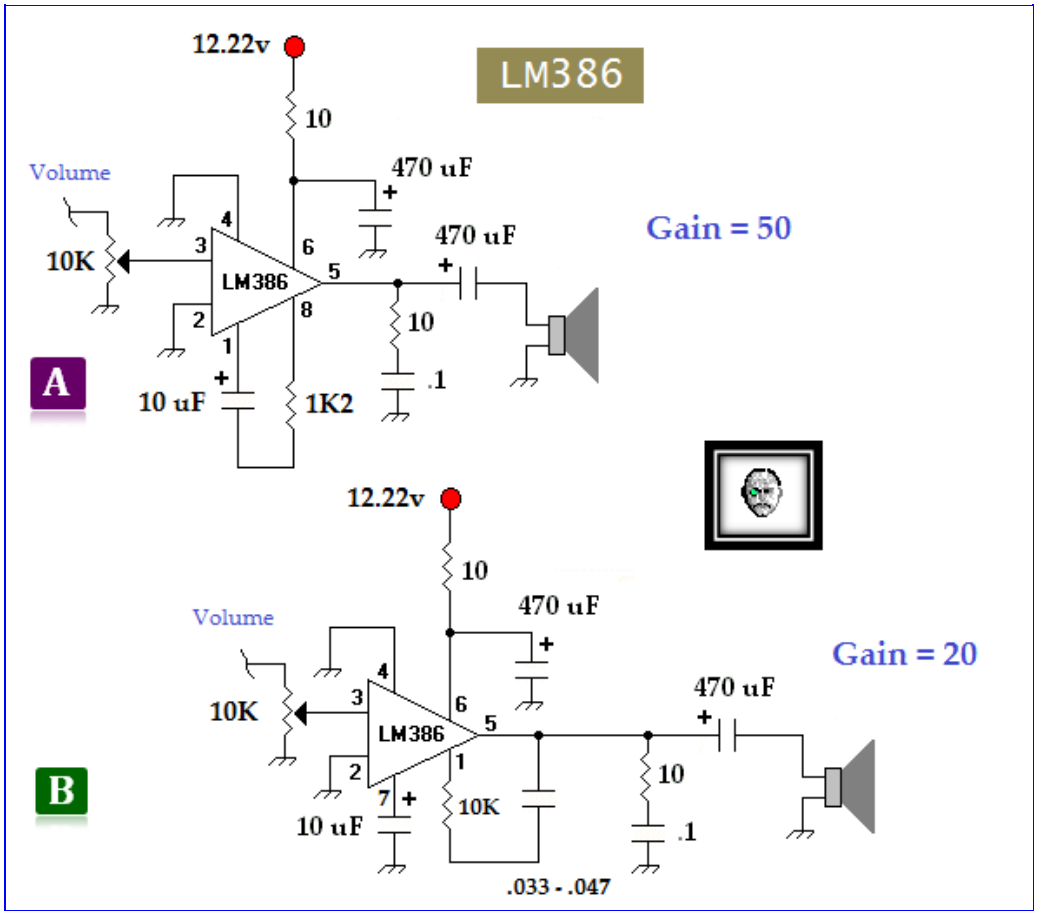
Arnie Coro talked about a "regenerodyne" receiver in Sound Byte *Audio3*. Very cool. Here's a link. Now this is radio!



Above — The Stupid-Simple regenerative receiver breadboard for 6 MHz. You can see the T68-6 — the red secondary windings are wound in the same direction as the 22 turns of primary. I started with 9 turns and unwound a link and tested sequentially until I had the right amount of feedback for AM. I just used normal hook up wire for the secondary winding. The white colored cable goes to the audio amp. This is a prototype experimental layout — a regenerative receiver should have short connections around the tank circuit and be in metal box for best results.

On some stations, my little 1 RF + 1 AF stage receiver sounded better than my superheterodyne receivers. The 2N3904 is just barely turned on — I determined that a base bias of 0.66-0.69v provided maximal sensitivity. The 150 ohm emitter resistor can be a 500 ohm pot and used to fine tune the regeneration. 150 ohms gave the best compromise gain and feedback + current for the 2N3904.

Audio Amplifier



Above — I chose an LM386-N for my AF power amplifier. The LM386 exhibits less peak signal distortion when run in the low-power (X20) gain mode and a higher VCC such as 12 volts. My receiver used the schematic denoted B — a 10K and 0.033 to 0.047 uF RC network is used to reduce the amplifier high frequency response. Click for a sound byte of me tuning around 6 MHz with a 10K + 0.047 uF RC network between pins 1 and 5. Figure B is my favorite way to use the LM386 and comes right off the National Semiconductor LM386 data sheet. Look for this data sheet with your favorite search engine.

Because I have a big antenna, Figure B provides adequate volume to a speaker. Connecting pins 1 and 8 via a 10 uF capacitor bypasses some emitter resistance and gives X200 gain. A resistor in series with the capacitor pin 1 and 8 will reduce the gain. Figure A shows a gain = 50 configuration. You'll have to choose the LM386 set gain to suit your particular regenerative receiver, however, the greater the gain setting, the greater the chance of distortion, unwanted noise and audio feedback.

I generally build my audio power amps around op amps or discrete transistors, but the LM386 exalts this web page's theme. Distortion in all these small power amplifiers is dependent on input signal amplitude as much as anything else.

LM386 Motorboaters

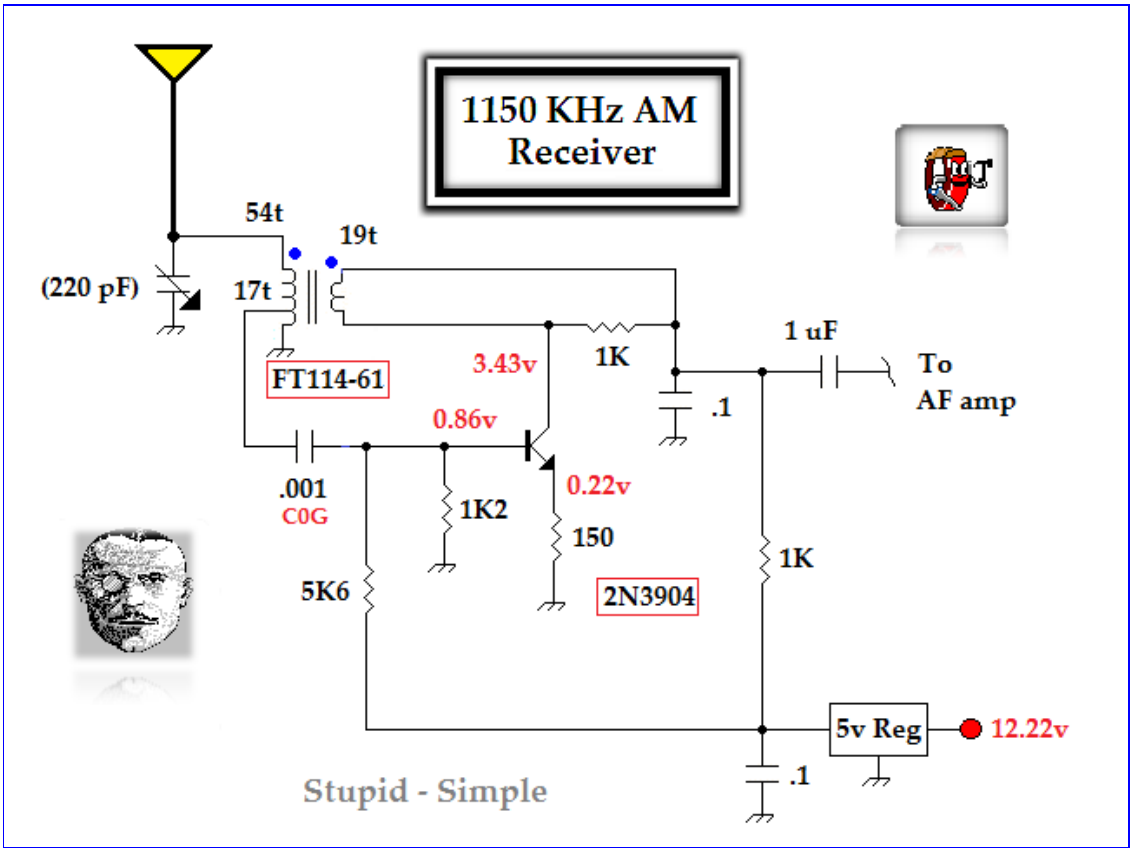
Expect AF feedback motorboating via your DC power lines as you increase the AF gain pot in *many* LM386 circuits. If this happens, try better bypass and decoupling on the LM386 power line: 100 to 470 µF shunt bypass on the DC line + a 10-22 series decoupling resistor, plus 100 - 470 µF shunt bypass on pin 6. Click for 1 example. ypically 470 µF bypass is required if you suffer 60 hertz hum. You may model your simple RC filter with application E on the javascript applet page to see the 3 dB cut off frequency of your particular DC line low-pass filter.

Stupid—Simple Notes

The Stupid—Simple circuit really needs the adjustable 10K regeneration control if you wish to tune both AM and CW. The number of turns on the feedback winding varies with factors including transistor beta, how you wind the primary and secondary windings (greatly affects the coupling between the primary and secondary windings) and whether you want AM, or CW reception — or both. Experiment with the number of turns to figure it out.

You can try "matching" the tank circuit to your antenna by decreasing the 470 pF cap to as low as 68 pF. This will affect the tuning capacitor range. For an air variable tuning capacitor, use anything you can find. Consider connecting fixed parallel and/or series capacitors to reach or limit the desired capacitance. Many good examples are published on the web.

Some builders float the tuning capacitor across the inductor so the cap is ungrounded. The stator (body) of the capacitor should be connected to the circuit ground to help minimize the effects of hand capacitance. A grounded metal case further helps.



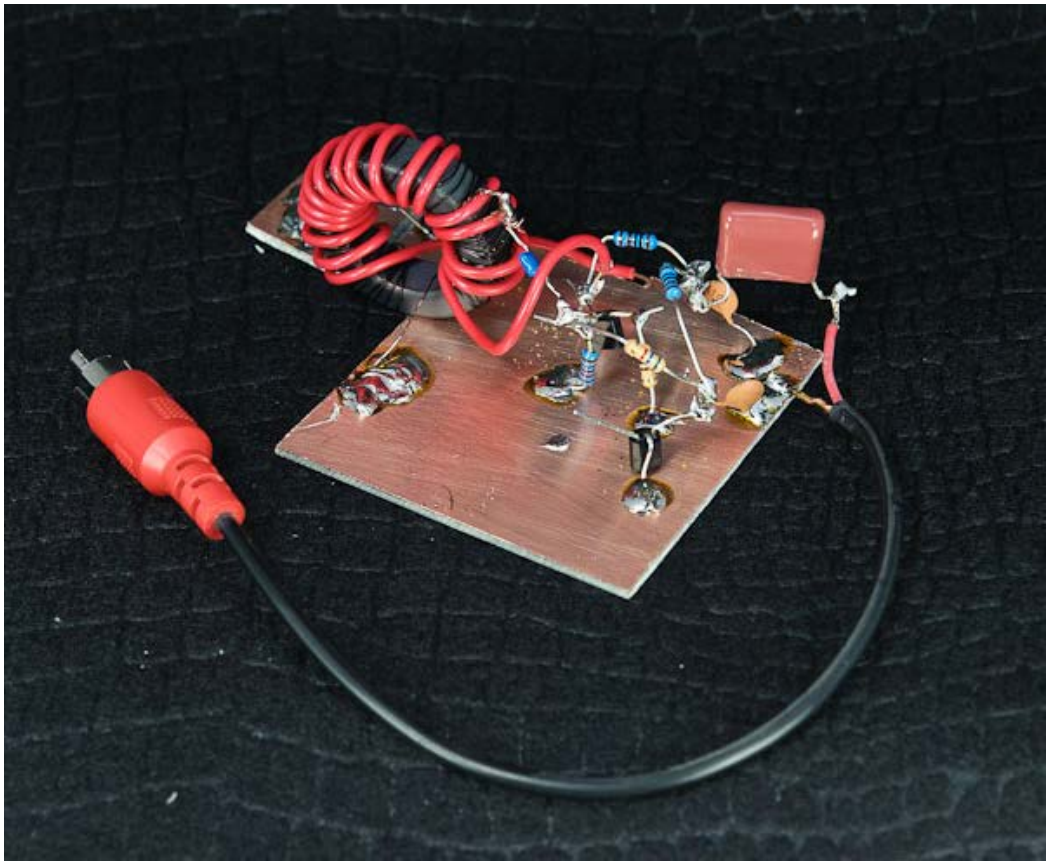
Above — My popcorn regenerative design "The Stupid-Simple" set up for broadcast band radio at 1150 KHz.

A reader from Brazil enquired about putting the Simple Stupid on MW. I had some time for a couple of experiments but only wanted 1 frequency — 1150 KHz, the local 10 KW sports radio station. Using 28 gauge wire, I wound 54 turns (about 230 uH) on a A FT-114-61. Most builders won't have this toroid, however ferrite rods from AM radios are plentiful and a great substitute.

This design relaxed the regeneration to improve audio quality (no whining or hissing). The bias and feedback loop were wound for the best sounding audio. For example, at the bias shown, if you increased the 19 turn link to 21 turns, the bass response increases; decreasing to 17 turns reduces the bass response.

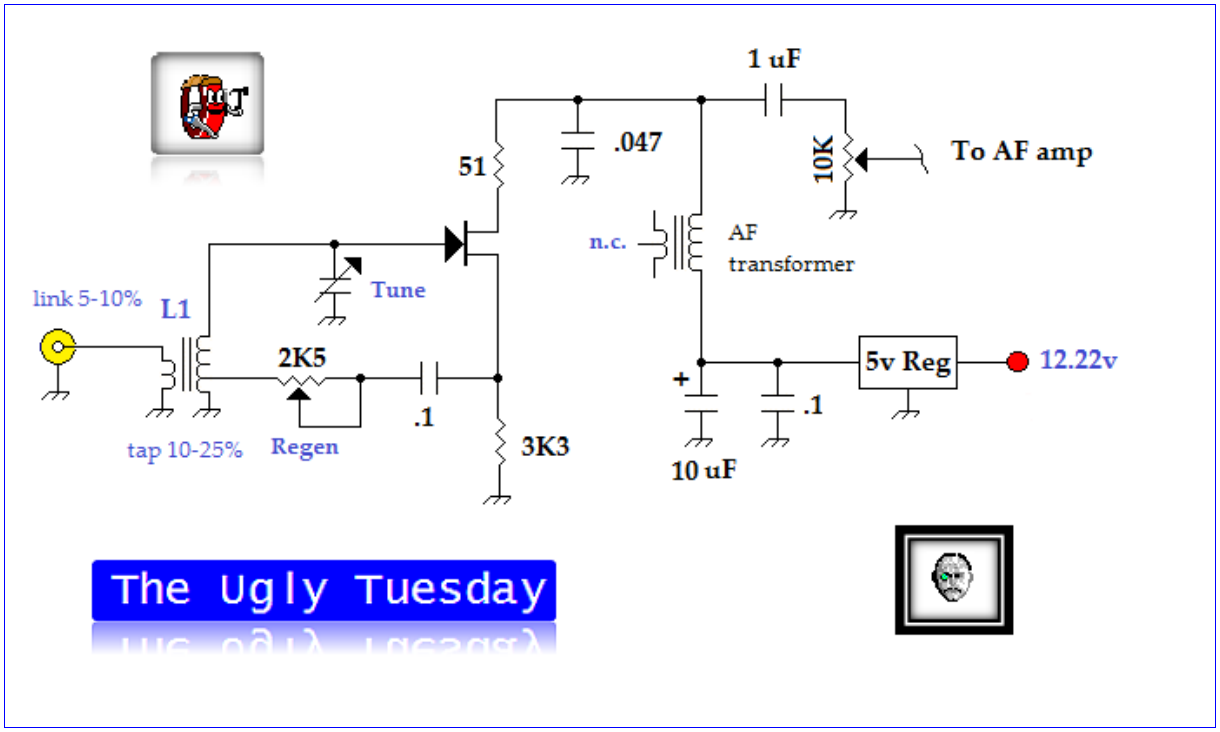
As a result of lowered positive feedback, the selectivity is down, however, a variable capacitor is needed to peak the station. After peak tuning, I removed and measured the air variable cap at ~ 200 pF and then substituted a 220 pF fixed capacitor to simplify things.

For audio, I used a bench AF power amp into a speaker. It sounds nice for 1 transistor. Audio sample



Above — The breadboard of the 1 channel receiver for my workshop - I'll use it to keep track of the Canadian Football League statistics. My test antenna was a long piece of outside wire. Red hook-up wire forms the 19 turn feedback winding.

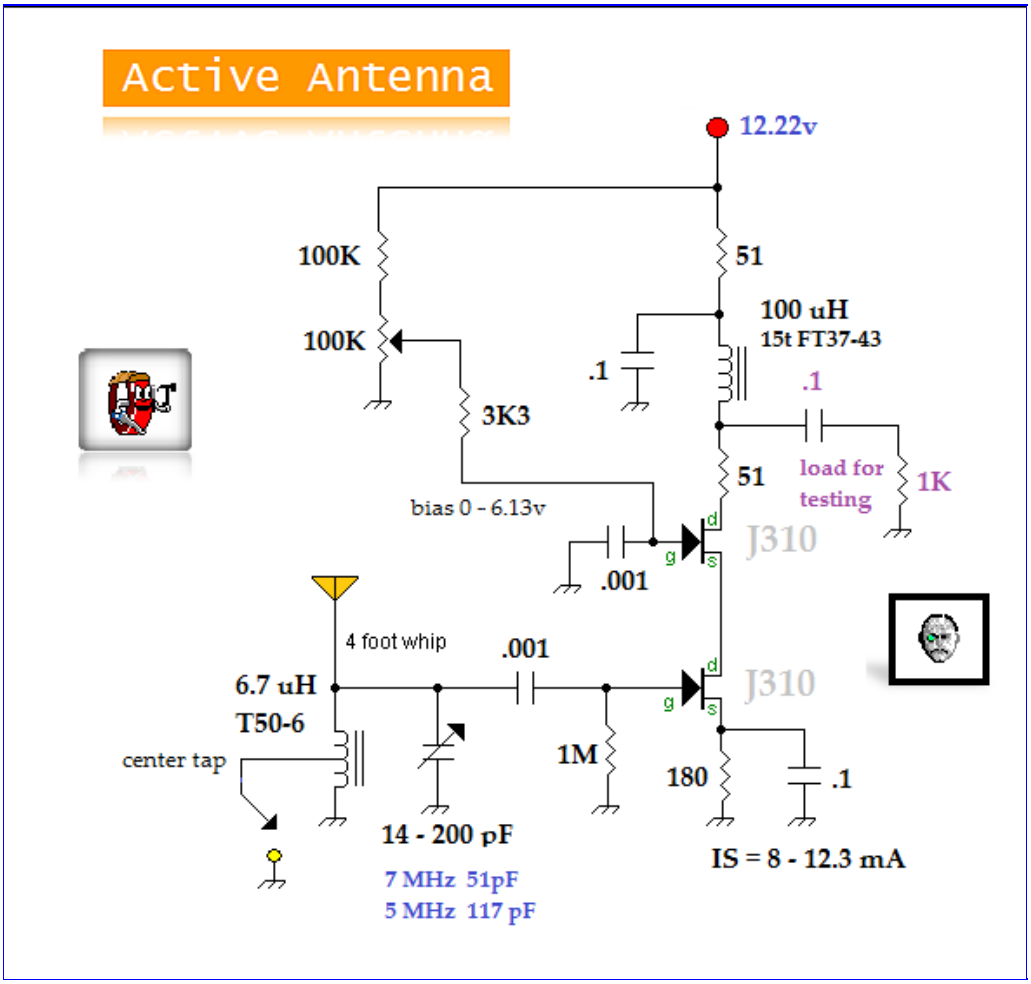
For cities with multiple AM stations (AM stations are dying out in Canada), you'll have to add more regeneration and probably move to a better design. This radio is simple, but not extraordinary.



Above — Another regenerative design that tuned AM, SSB and CW from 5-11 MHz with different toroid coils wound on a T50-6. It's based on a favorite design by Wes, W7ZOI. I suggest tapping L1 at 10 - 25% of the total number of turns. The secondary link for the antenna connection depends on the impedance of the antenna, but 5 - 10% of the total number of L1 primary turns worked well at my QTH. Please experiment with the secondary link to determine the optimal coupling to your antenna. My L1 inductance ranged from 1.5 to 5.6 uH. You may have to add a fixed capacitor in parallel with your air variable capacitor when using low inductance coils such as 1.5 uH.

The 51 ohm resistor suppresses UHF parasitic oscillations. The AF transformer is a transistor radio output (1000 : 8 ohm) junk box special and serves as an RF choke. I tried various AF transformers harvested from old transistor radios in this slot and they all worked fine. Nothing's really critical on this receiver — that's why I like it. Truly junk box radio.

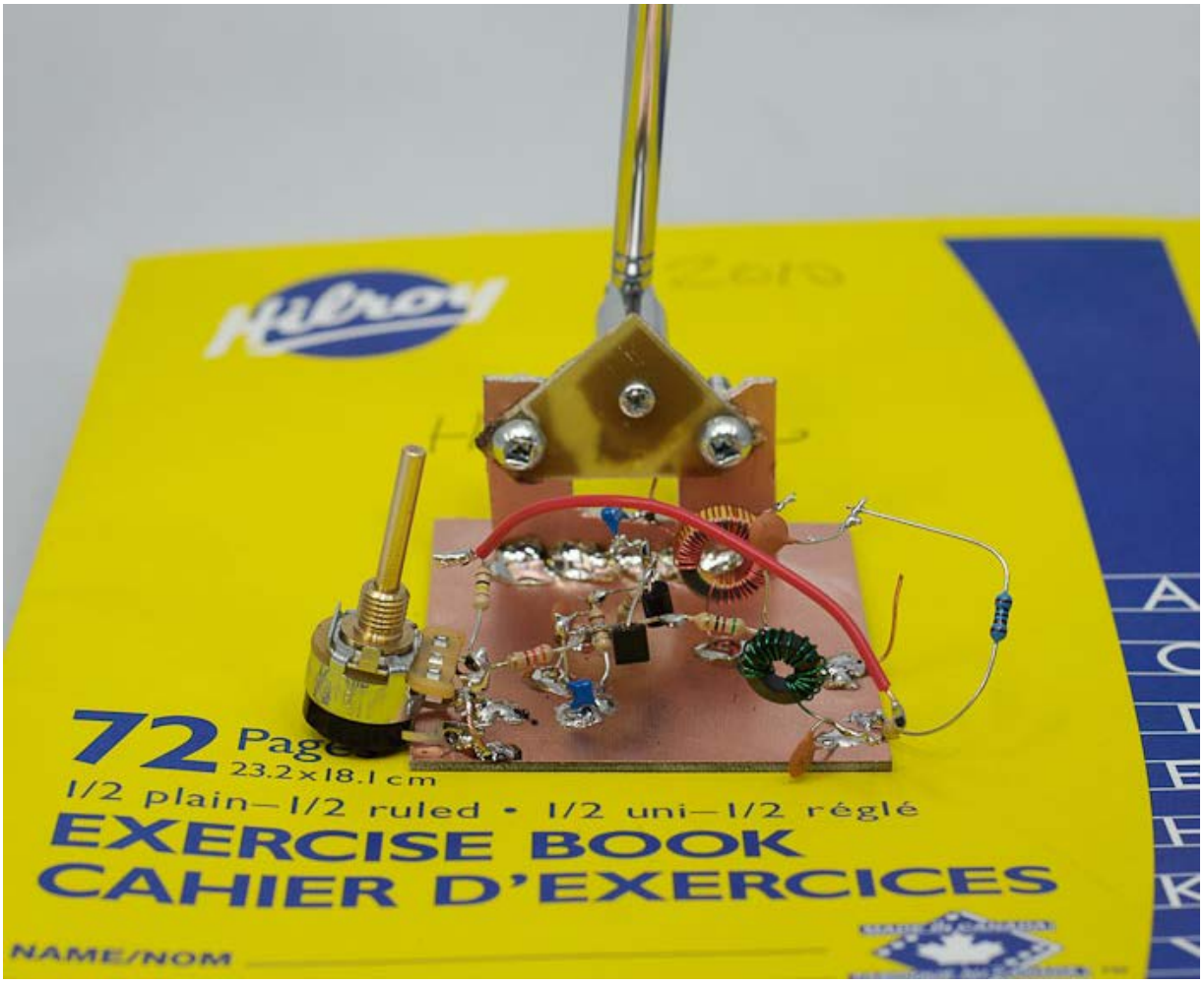
Simple Active Antenna Experiments



Above — A voltage probe or active antenna using a telescopic whip. It's been awhile since I built one in keeping with a minimalistic circuit theme. I tested this VPA from 5 to 14 MHz. The center tap on the coil allows the peaking at ~10 MHz and higher. The L value is non-critical; choose a value that will work with your tuning variable capacitor or varactor. The L - C values can be *roughly* determined from a chart like this, or just do the math (XL = XC at the desired frequency). Account for stray inductance. If you wish to perform return loss measurements on this circuit, you'll have to short the 6.7 uH inductor as the whip antenna can tune in RF from the RF signal generator used for the return loss bridge.

Without the 1K load, the circuit will oscillate. I thought about some ways to match the output transformer to a regenerative receiver tank circuit. The 100 uH drain choke could be replaced by a (bifilar) 2:1 transmission line transformer or two. Transformer experiments this Winter clearly illustrated the superior coupling of transmission line transformers and mandates using them over conventional transformers whenever possible.

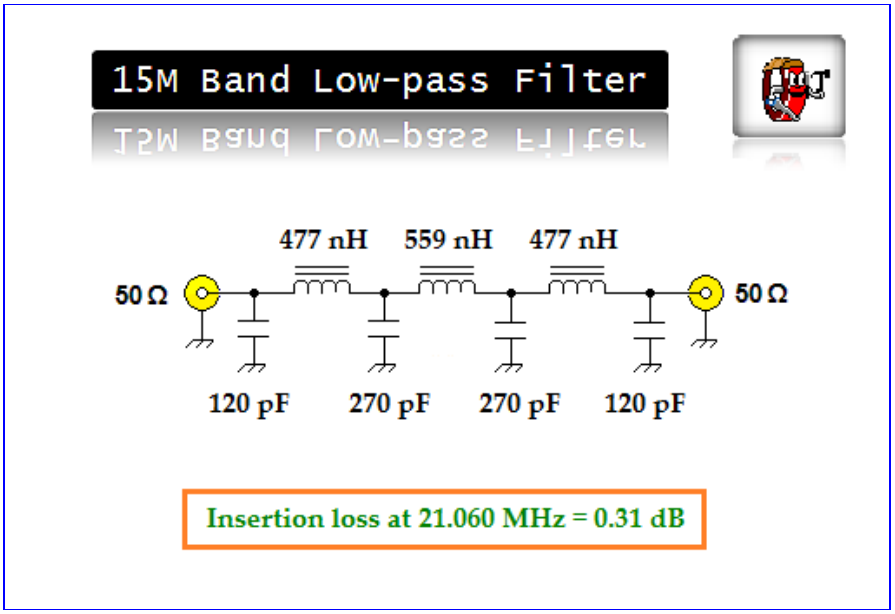
Using a conventional transformer with a shunt resistor across the transformer would also work, but the resistor reduces gain. I built and tested this output circuit with an 8:1 transformation using two 2:1 transmission line transformers. The output impedance at the JFET drain is somewhere around 4300-4500 ohms at 7-14 MHz. The transformed output impedance is somewhere around 250-330 ohms at 7- 14 MHz. Connecting the VPA output to a tap in the regenerative main tuning inductor might work — being careful not to load down the regenerative tank coil.



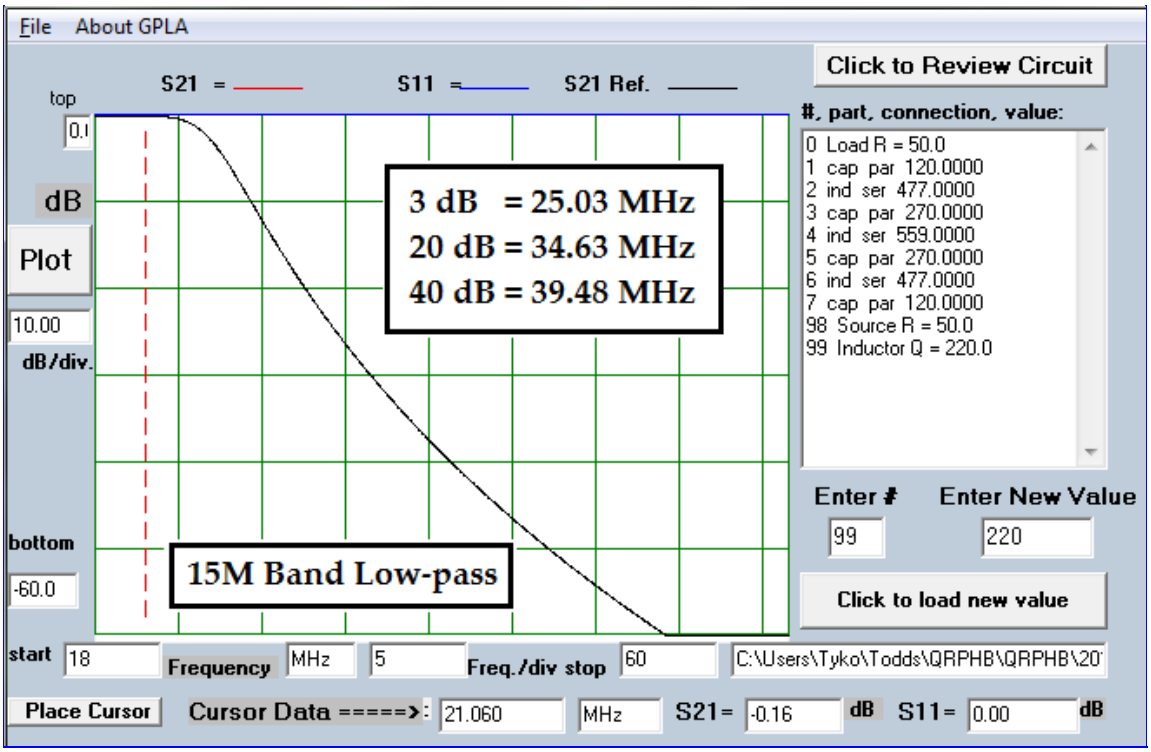
Above — The VPA breadboard on my latest notebook. The 100K pot sets the stage gain. While simple, it works okay.

Above — the VPA built March 12, 2011 (the 14-200 pF air variable cap is not shown). It took about 45 minutes to design, build and test it. The 1K load used for testing is the blue resistor to the extreme right.

Low-pass Filter for 21 MHz



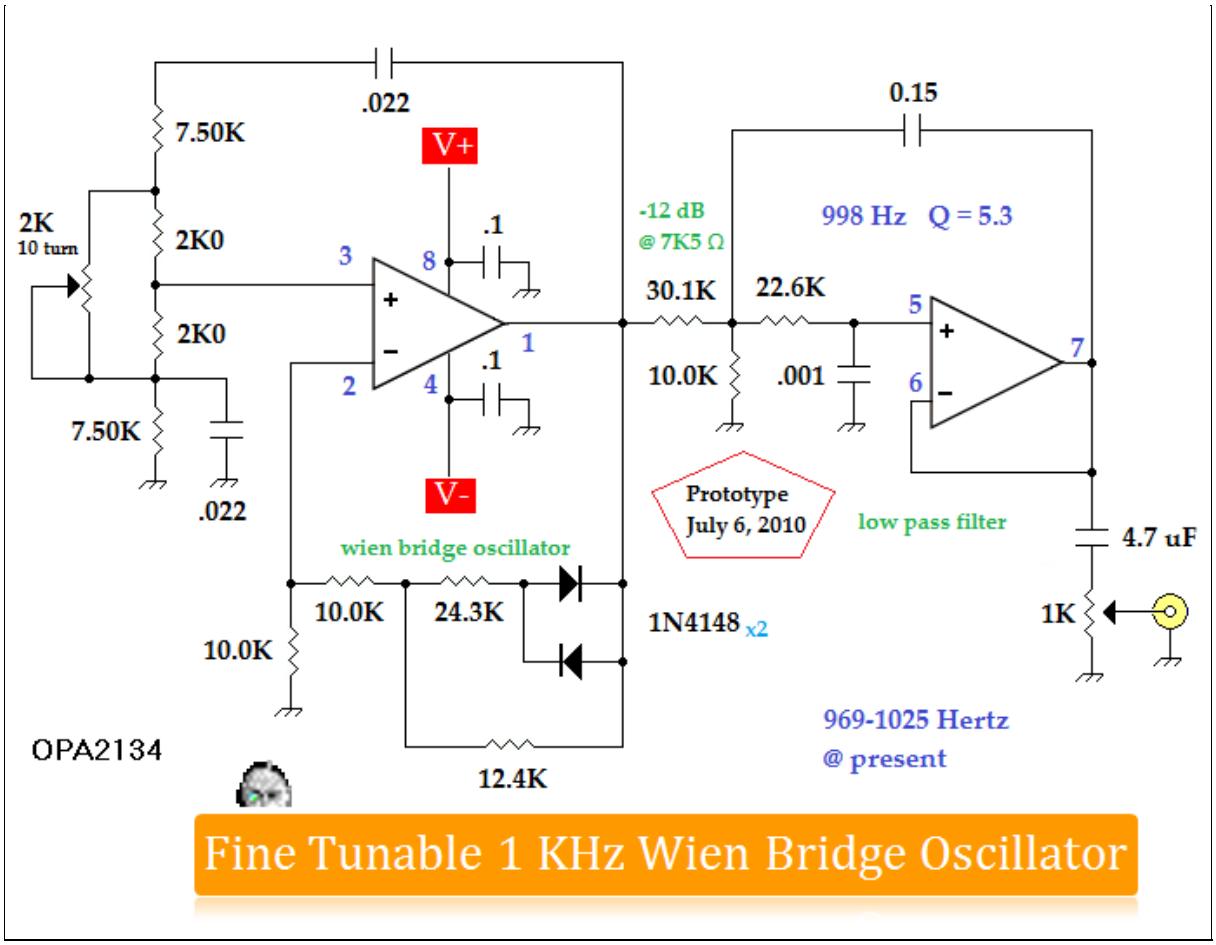
Above — A 7 element Chebyshev low-pass filter for the 15 Meters Ham band (fCo = 25.03 MHz to allow the use of standard value capacitors). A builder requested a band-pass filter design for his 15M band receiver. In order to accurately test my design, I decided to make a permanent, low-pass filter module to follow my signal generator. 15 Meters is a favorite Ham band, so I'm certain to use it in the future.



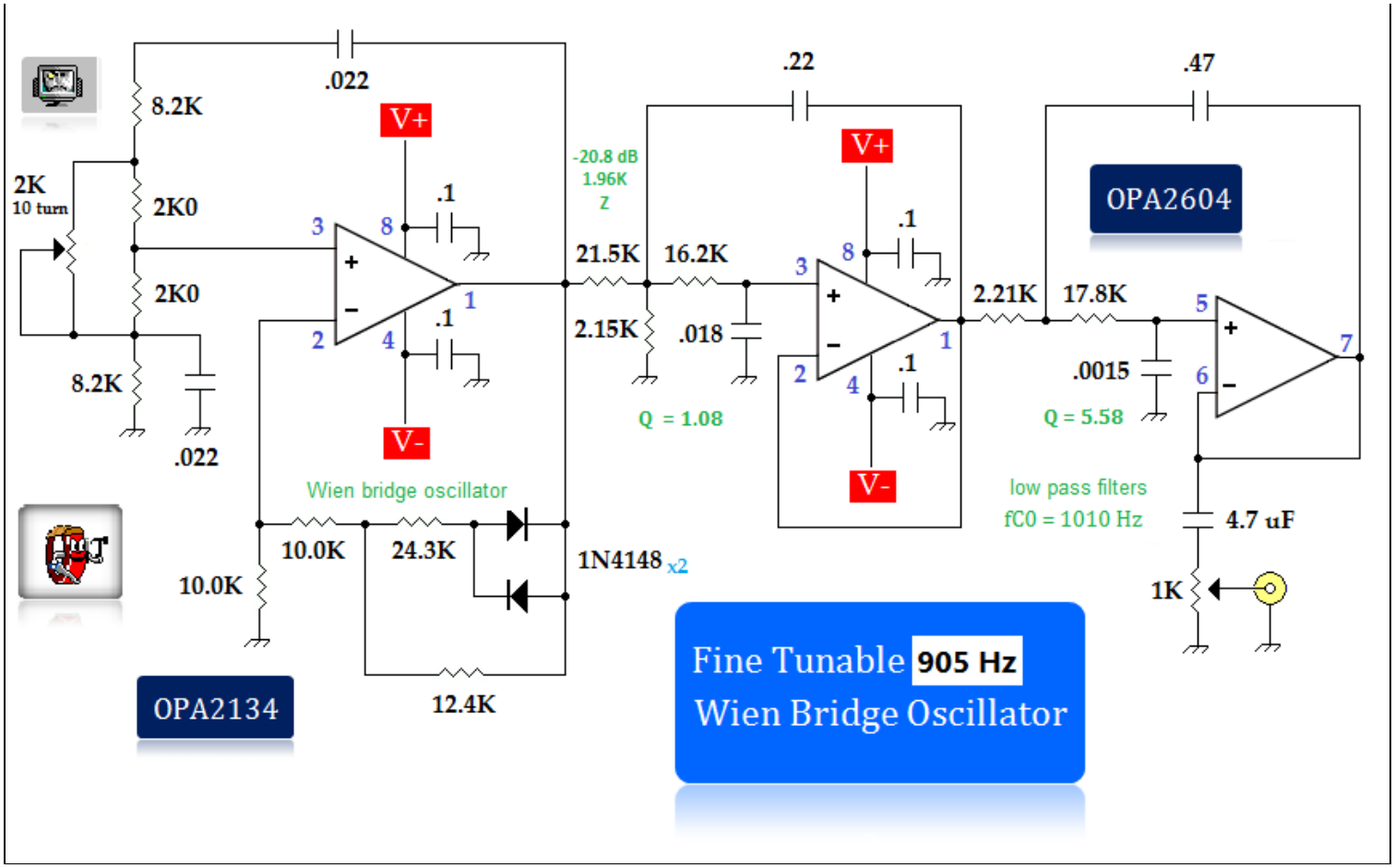
Above — A GPLA plot of the filter. The frequency cut-off at -3, -20 and -40 dB are shown. Perhaps this filter is overkill, but I had all the parts on hand and love a serious low-pass filter. Click for the bread board photo.

Fine-Tuneable 1 KHz Wein Bridge Oscillator

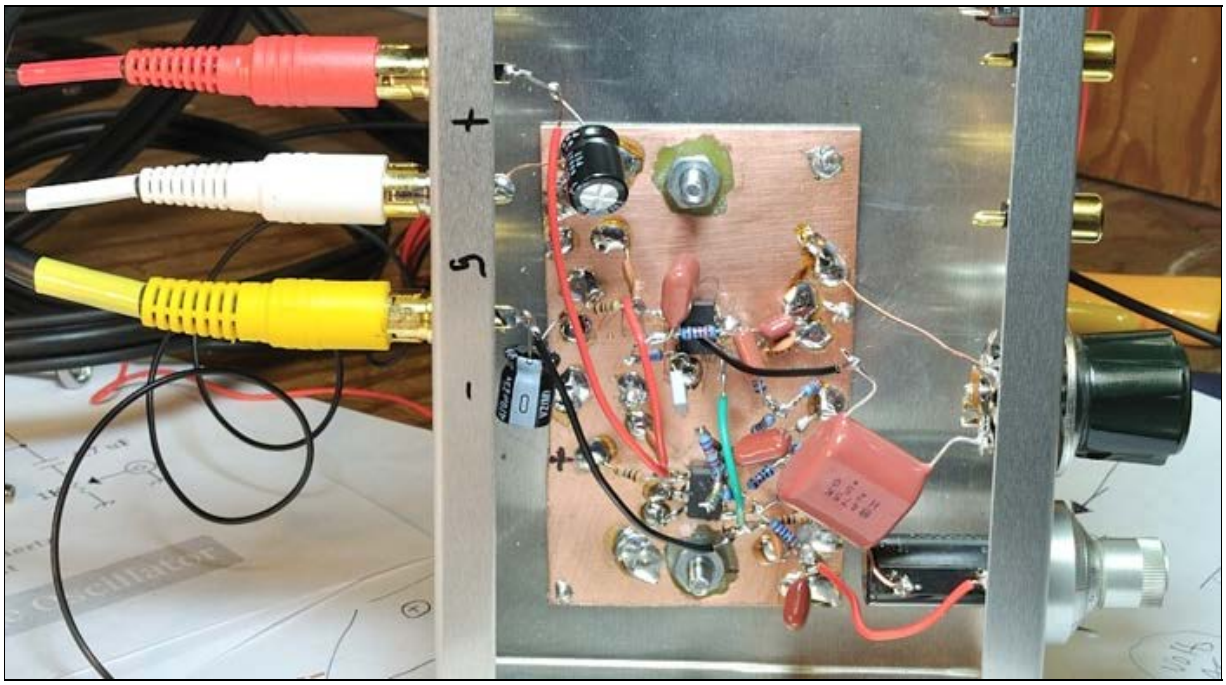
In 2010, I wanted a fine-tuneable Wien bridge oscillator to drive a notch filter in an AF distortion analyzer. Ken Kuhn drew me up a schematic on his coffee break and emailed it the same morning.



I applied anti-parallel diodes instead of the classic incandescent bulb for amplitude stabalization in the feedback loop; probably a mistake leading to higher distortion. Ken's fine-tuning circuit works perfectly. I matched the 7K5 + 2K0 + 0.22 uF components on each filter half. 1% parts go in this circuit.



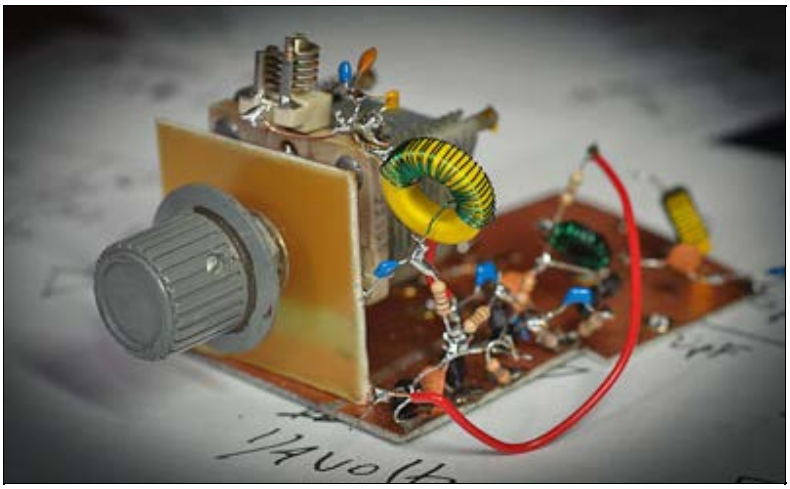
A version for 905 Hertz built with crazy expensive op-amps. I chose 905 Hz to match my notch filter frequency.



My breadboard.

QRP Modules 2011

Introduction

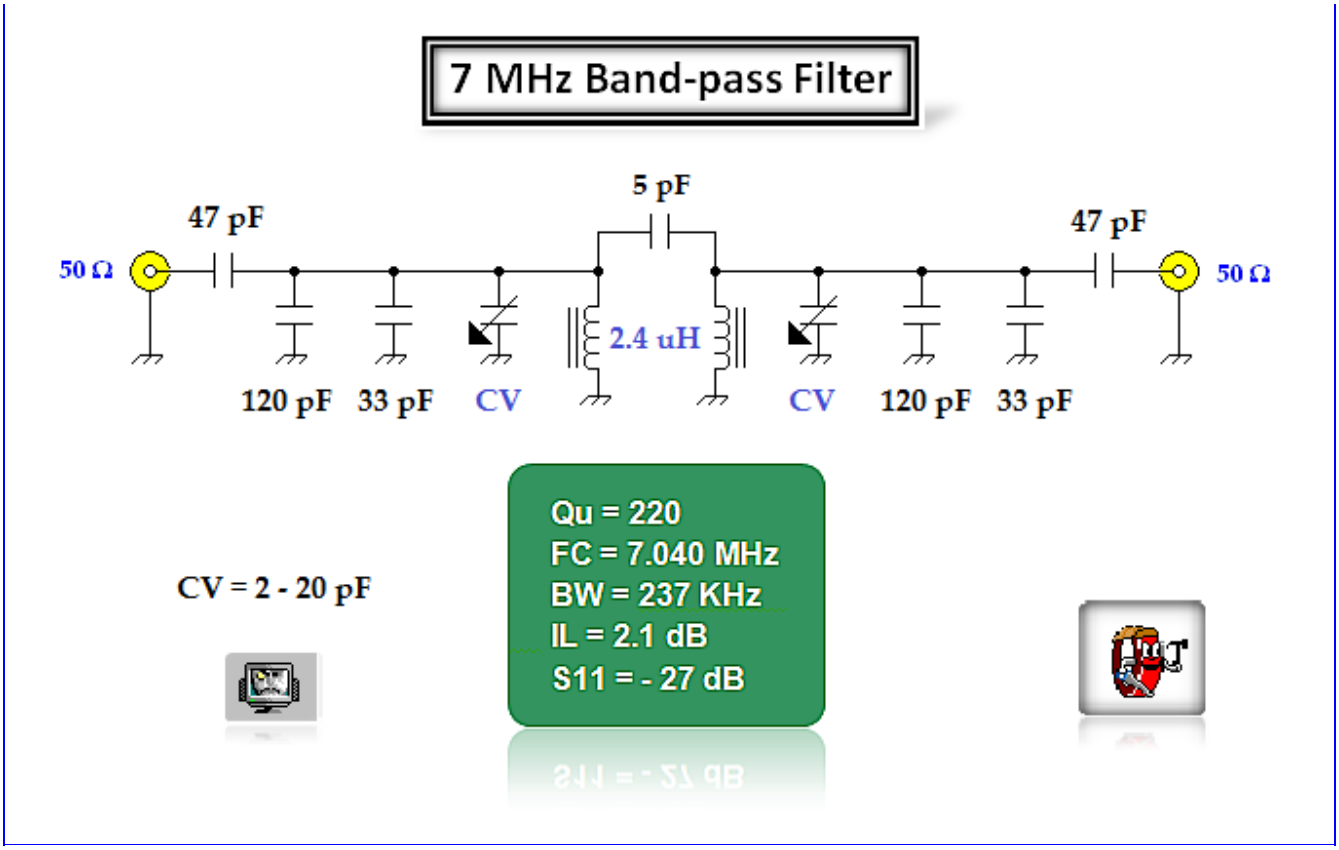


As experimenters, we rebuild core circuits over time. I decided to increase my collection of stock modular circuits to avoid re-inventing the wheel. This web page serves as a module repository for the website.

Since our needs differ, I've shared these circuits more for interest sake and really not as schematics to copy. All modules were carefully built and tested.

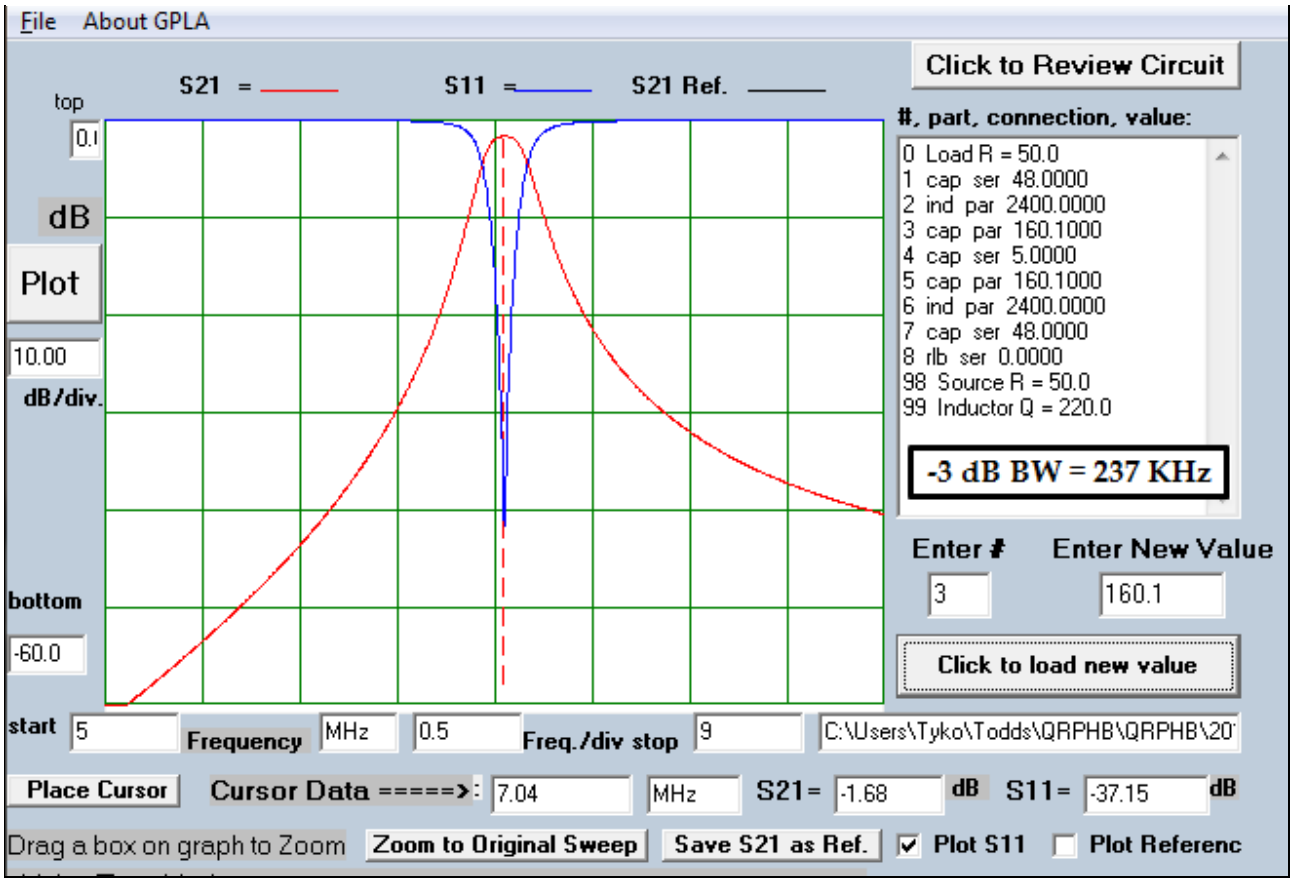
1 great virtue of the metal encased module is strong shielding. RF modules use a 50 ohm port impedance and BNC connectors. RCA jacks interface the AF modules.

40 Meter Band-pass Filter

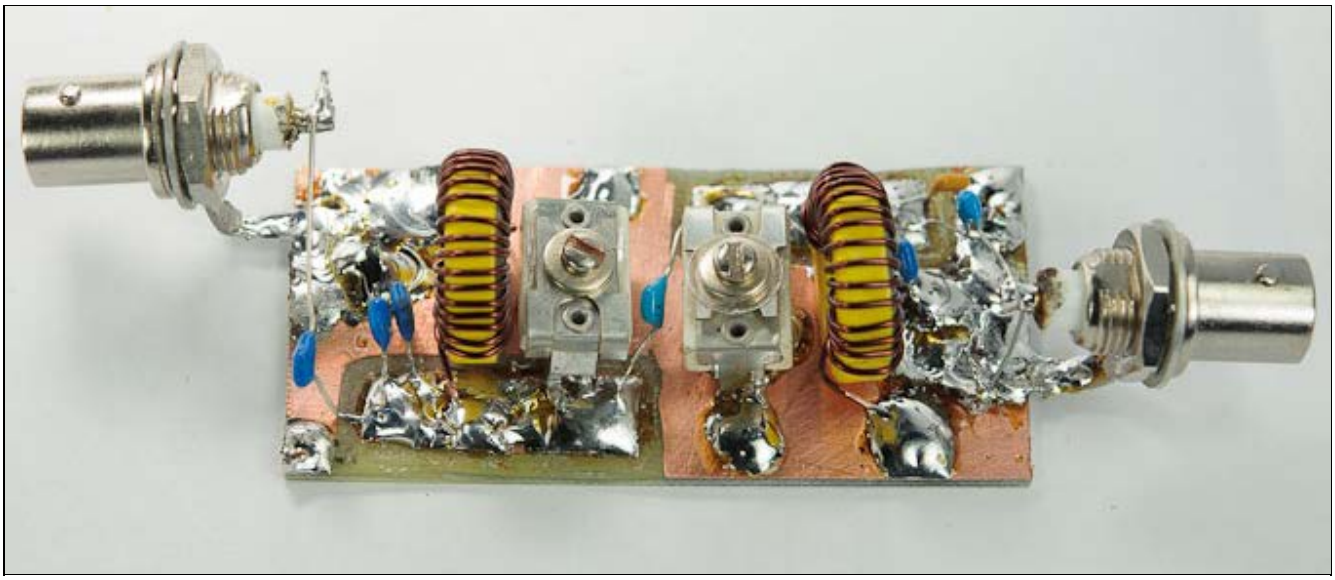


Above — A 40 Meter Ham band double-tuned band-pass filter. I designed this circuit using 2 programs that came with EMRFD and describe the process on this web page. The 2.4 uH measured coils were wound using #22 AWG wire on T68-6 powdered iron toroids and all fixed caps were ceramic C0G type. I centered my filter at 7.040 MHz. You should be able to peak it anywhere on the 40 Meter CW sub-band by tweaking the variable capacitors.

I peaked the trimmer capacitor while looking at the peak-to-peak voltage on a 50 ohm terminated oscilloscope. The filter input was connected to a 7.040 MHz signal generator with a 30 dB return loss, low harmonics (-55 dBc) and 50 ohm cables.



Above — A simulation of my filter design in in GPLA08. The calculated IL was 1.68 dB, I measured the IL at 2.1 dB The calculated return loss or S11 was 37.2 dB; I measured 27 dB. A good filter.

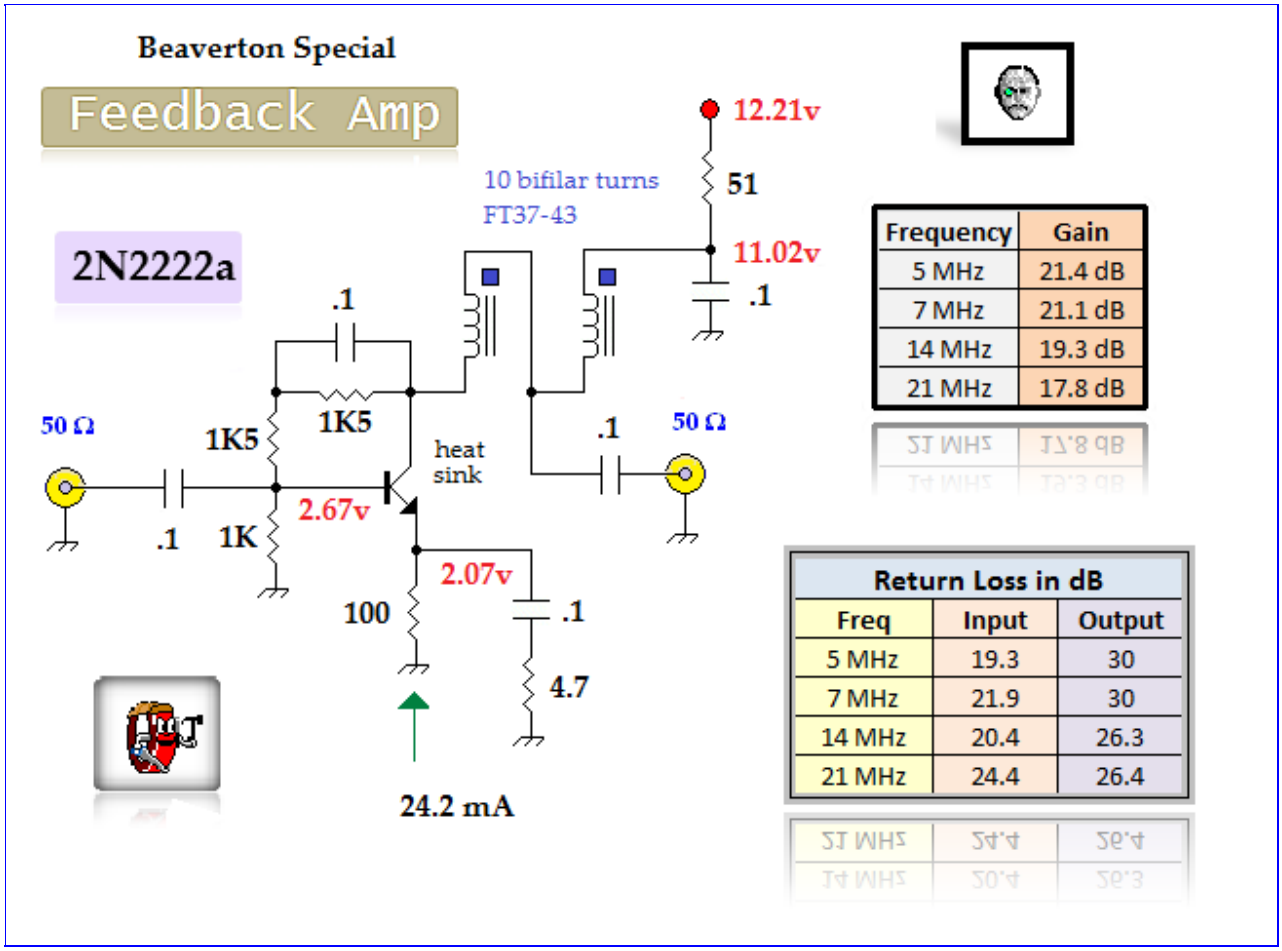


Above — the 40 meter double tuned band-pass filter breadboard with temporary BNC connectors and series caps. Since this filter will serve as my main front-end filter for all future 40M band receiver bench design, I blinged out and put in big toroids and high Q, air-variable trimmer capacitors. While I could have just use a single 150 pF tank capacitor and a wide range trimmer cap such as common, ceramic 10-70 pF, the small range, high Q trimmer capacitors offer better performance and fine tuning.

Click for a spectrum analyzer +tracking generator sweep where the center frequency = 7.040 MHz. Graticules: Horizontal = 1 MHz per division, Vertical = 10 dB per division. You can see why it tunes so sharply.

After testing the bread-board, I removed the temporary BNC connectors and series caps, I stuck it in a Hammond box and wired in permanent, short leaded 47 pF capacitors. Final testing in the sealed box varied minimally from the open bread board. This board looks especially ugly because it held a previous filter and contained lots of remnant solder.

Broadband Feedback Amp



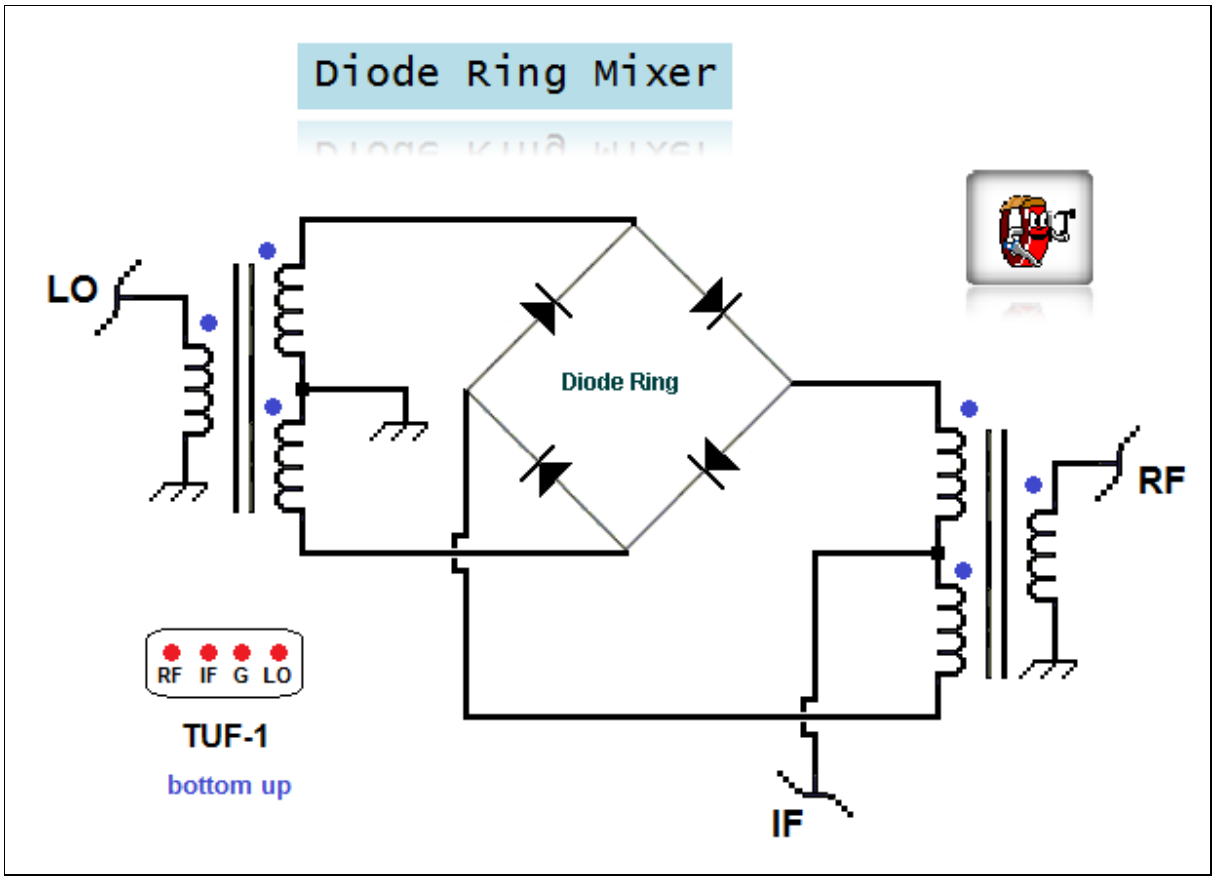
Above — A "Beaverton Special" feedback amp with analysis. As experimenters, we often need a go-to, broadband 50 ohms input and output RF amplifier. This is it!

Popularized by Wes, W7ZOI and Doug DeMaw, W1FB, this amp has stood the test of time and fits perfectly into the 50 ohm module concept. A bevy of transistors were tried — a 2N4401, 2N5179, 2N3904, 2N3866 or 2N5109 all worked fine. For the greatest return loss and signal handling possible, current over 21 mA is required and thus a 2N3904 isn't the best choice. Collector current = heat, so heat sink the BJT as appropriate.

I found that a 2N2222a biased with over 22 mA emitter current gave a stellar output return loss and low distortion. Within reason, for different transistors, keep the bias and feedback resistors constant and change the emitter resistor (100 ohms in my amp) to set the current you want or need.

Many builders follow this amp with an attenuator pad to preserve the input return loss.

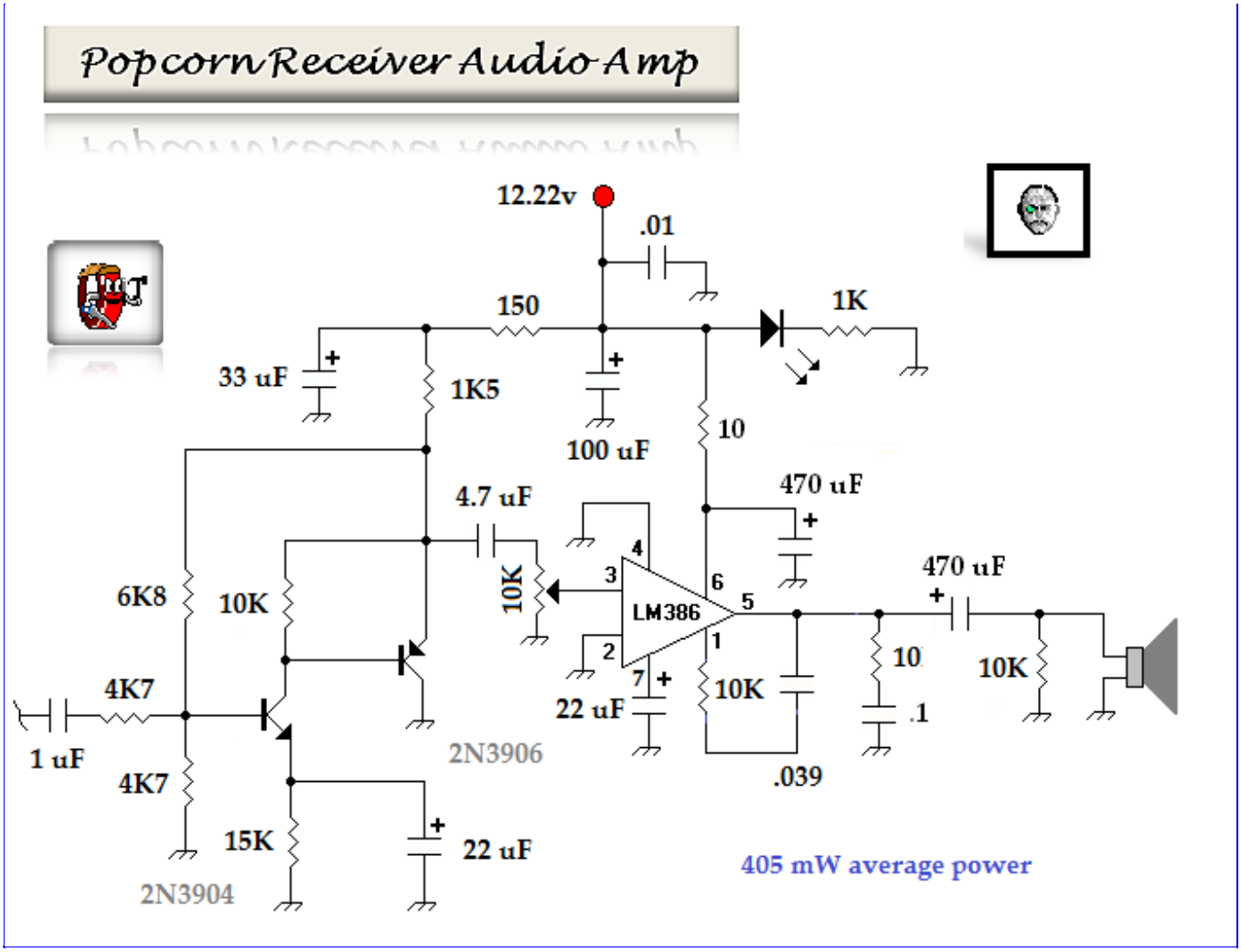
Diode Ring Mixer



Above — A Minicircuits TUF-1 diode ring mixer was used in standard configuration. 7 dBm LO drive.

Above — The DRM module. It's hard to photograph inside a solder laden chassis. Connections are short.

Popcorn Audio Frequency Power Amplifiers



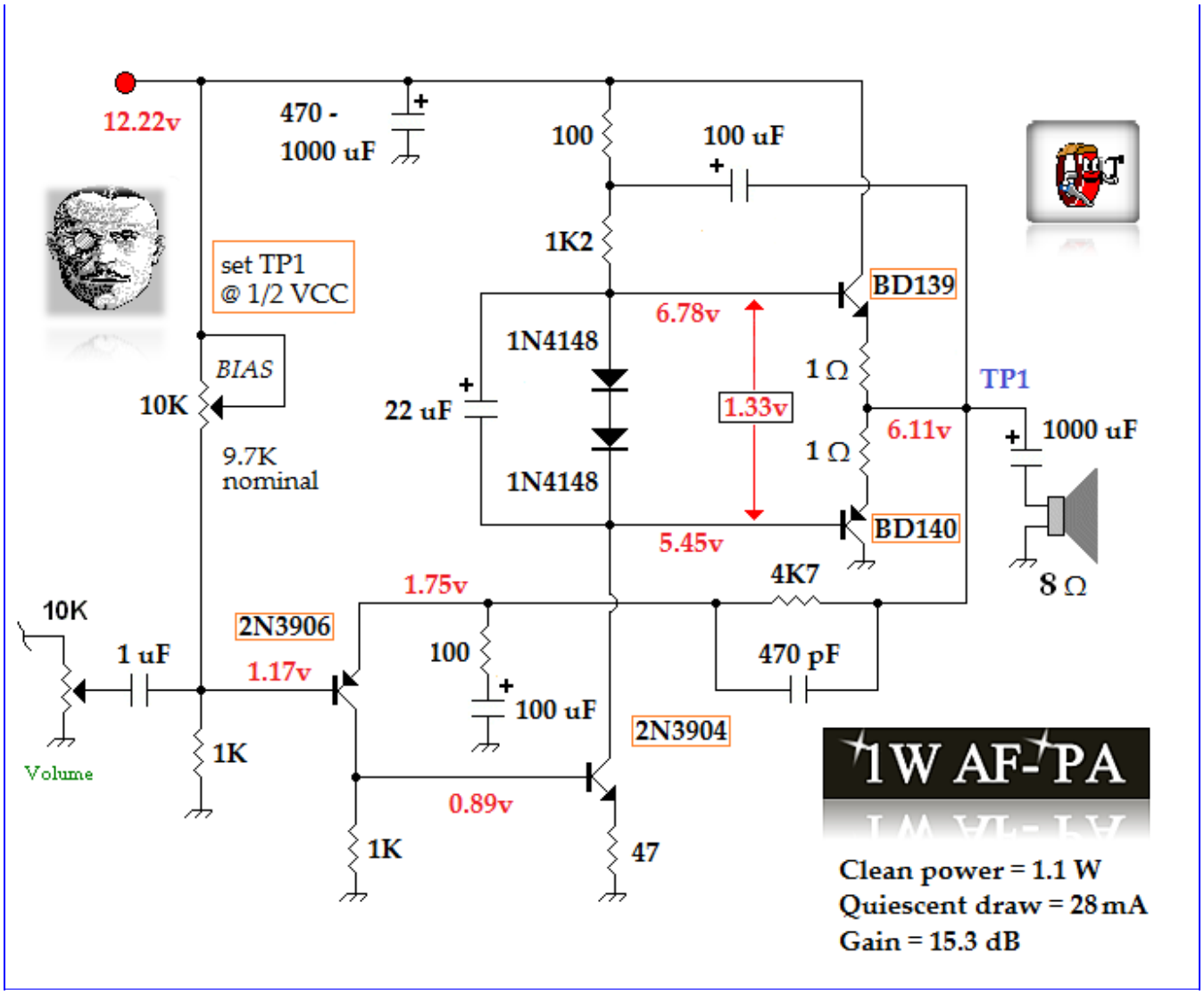
Above — Popcorn receiver audio power amp. I wanted a simple audio stage for testing popcorn receivers (to follow a high output impedance preamp device). Completing this module means never having to build such an amp again. The voltage gain is provided by BJTs to keep the noise down, but the popcorn factor up. The preamp impressed me with its strong signal handling capacity via feedback and careful biasing. The NPN is center biased so that when its intentionally distorted during testing, the positive and negative halves of the AC waveform distort equally — it provides a nice, big, AC voltage swing. An LM386 in X20 gain mode with some bass boosting comprises a reasonable power amp section. The 10K resistor on the output discharges the 470 uF cap when no speaker is connected to avoid a loud pop. The 4K7 series input R can be lowered, or omitted for more sensitivity.

Some builders might employ the LM386 in a higher gain mode at the expense of fidelity, or just wire up a TDA7052. I think in popcorn circuits, what really matters is that you understand what you're doing and try to *design* rather than just copy the "usual circuits".

Above — The popcorn AF amp in a clear blue chassis. Phono jacks provided a connection for the input and output — they 're inexpensive and readily available. The DC supply is connected to uninsulated banana jacks on the rear; it's well decoupled (resistor) + bypassed (capacitor) to help stop parasitic AF feedback. This amp is pretty quiet, considering its junk box legacy.

Above — The project with the top cover removed. The board is secured by the ground wires connecting it to the pot, jacks and DC voltage posts. The input is on the right.

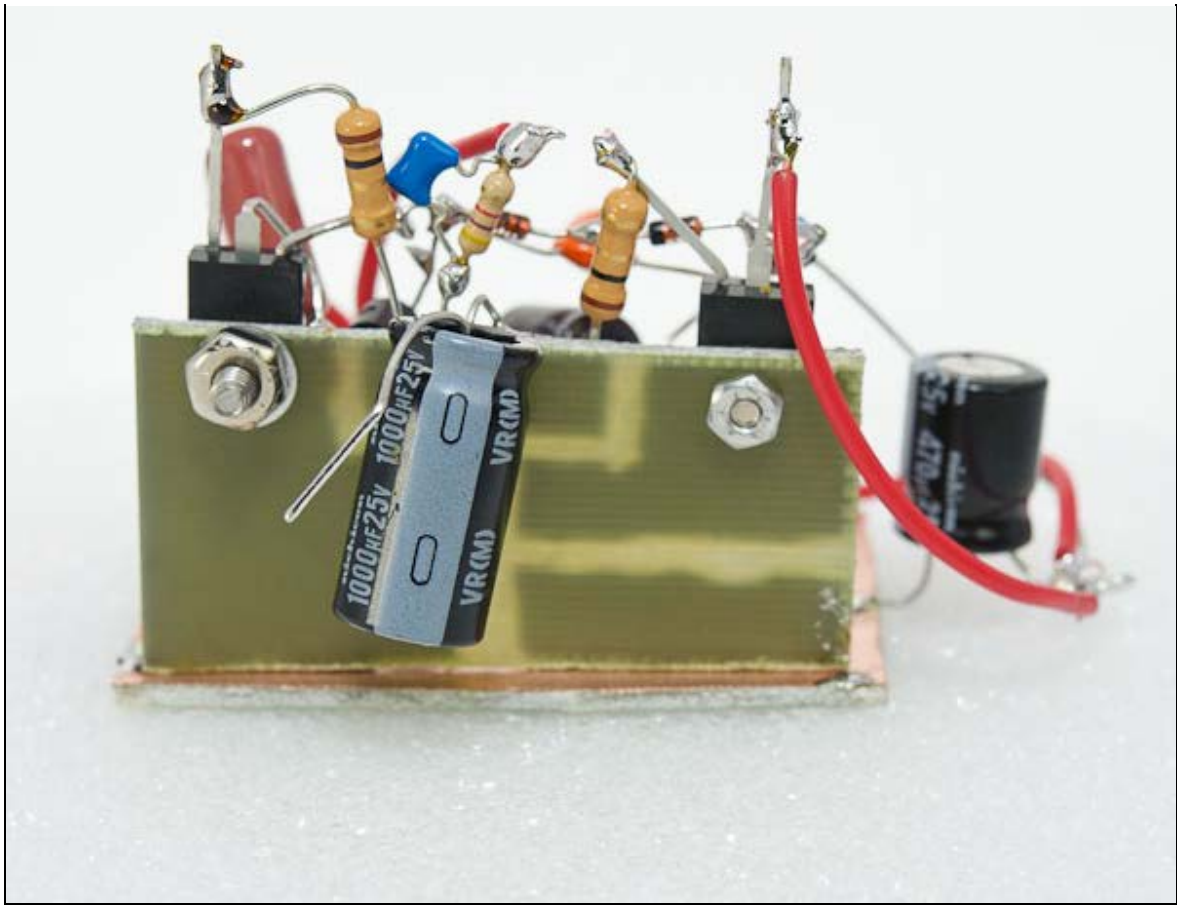
I like a relatively simple, lower gain AF amp on the bench for receiver development. You can use such an amp to decide on how much overall AF gain is needed, how you'll distribute it, and not have to deal with unwanted AF feedback.



Above — 1 watt popcorn audio power amplifier. Built around the BD139/140 complimentary pair - I achieved a clean I KHz sine wave at 1.1 Watts power after testing + tweaking my prototype design. I chose the familiar series diode pair to bias the power followers into Class A/B; an amplified diode (transistor level shifter) might be a better choice.

Bootstrapping the 2N3904 voltage amp pumps up the clean signal power capacity. The 2N3906 establishes the bias for the 2N3904 and the BD139-140 pair. Set the 10K bias pot so that the DC voltage at TP1 is 1/2 of the VCC. During testing, the AC voltage was centered perfectly between the DC rails and when pushed into clipping, the positive and negative AC waveform distorted nearly equally. Quiescent current = 28 mA ; not meant for a field-portable receiver.

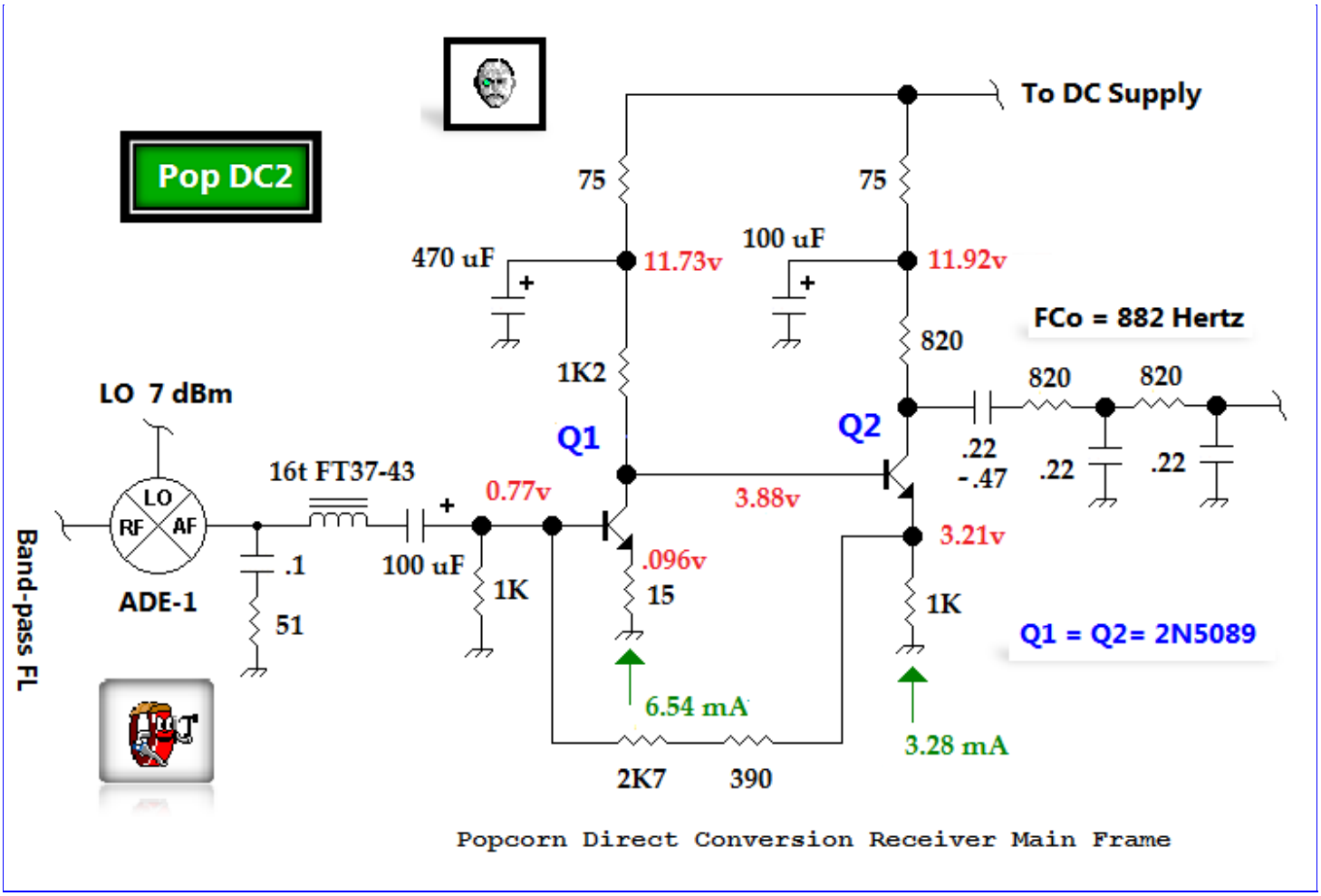
Click for a photo of my breadboard. Copper clad board serves as heat sinks for the power followers. The BD139 and BD140 make great complimentary transistors for audio frequency power amplifiers. With an Ft of 190 MHz, the BD139 can work okay as a driver or even the final in modest power QRP transmitters.



Above — 1 watt Audio PA (reverse view).

Pop DC2 — Popcorn Direct Conversion Receiver Main Frame

These circuits update the Popcorn DC receiver from 1998 and includes all components from the product detector through to the speaker, minus the VFO and band-pass filter.



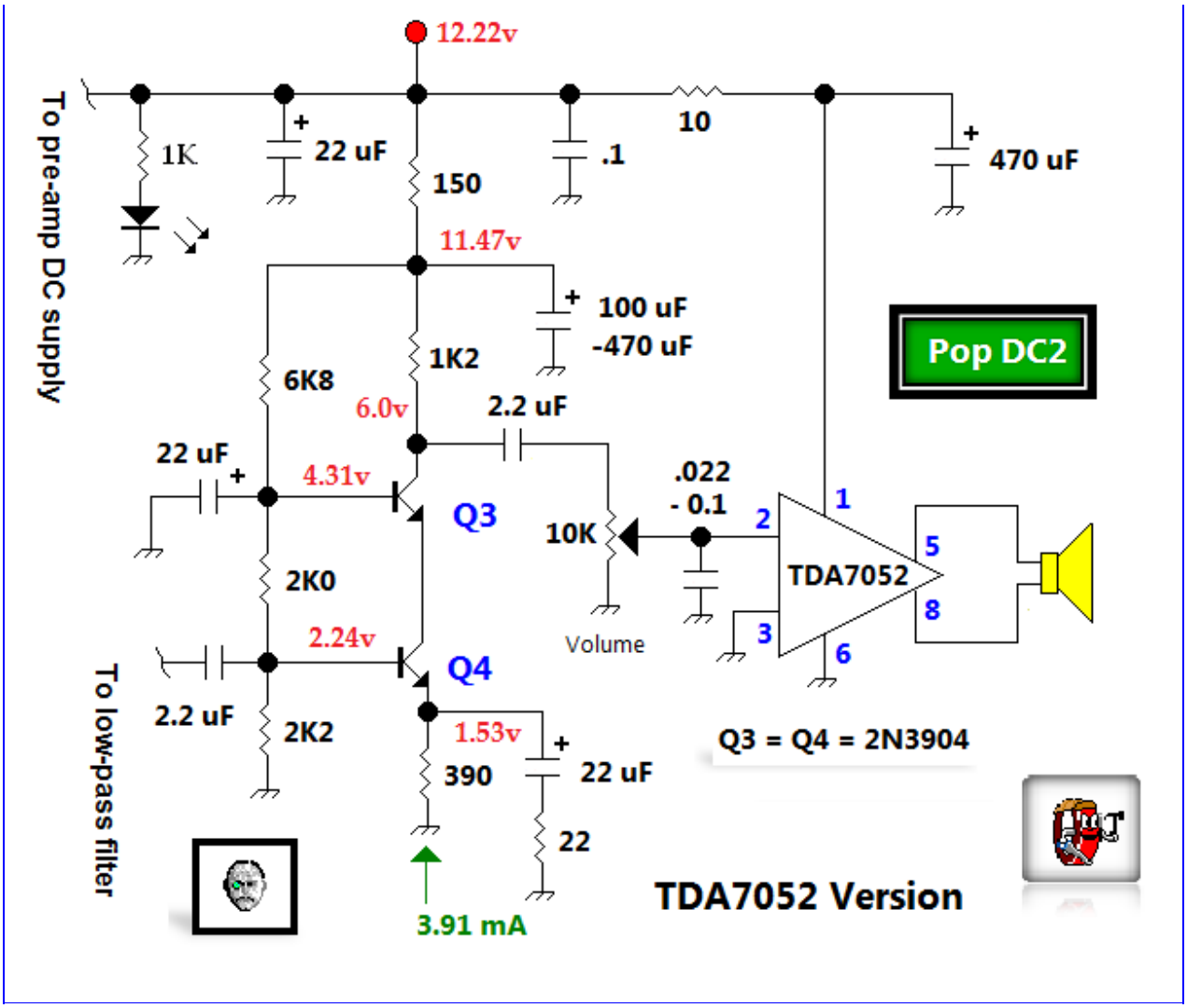
Above — The mixer and first audio preamplifier The 0.22 uF to 0.47uF cap connecting Q2's collector to the low-pass filter network exerts a high-pass response to remove low frequency noise and potentially any hum. I heard no hum, although a 470 uF filter capacitor on Q1 helps ensure that. Increase the 100 uF filter capacitor filtering Q2 if you hear hum or motor boating. The diode ring mixer exhibits AF that's hard to beat — very dynamic, vibrant and lively. I enjoyed the low microphonics with the double balance + a return loss of over 25 dB on all of 3 of its ports. Alternate photo.

I've read negative comments about my use of "those big filter capacitors" — 1 thing radiophiles can learn from audiophiles is that to adequately decouple and bypass means we need to stop fooling around with the usual 22 - 47 uF capacitors and *really* bypass. Viewing well designed AF amplifiers informs us so; these designers *really* filter their amplifiers from the DC supply. You can always increase the decoupling resistor value to allow use of a smaller capacitor value, however, we only have a single power supply at around 12 VDC, and I dislike giving up too much of it for DC filtering purposes. Do what ever amuses you.

Click for some analysis of the preamp. The MPSA18 went obsolete in 2011, so I chose the low-noise 2N5089 for Q1 and Q2.

The Popcorn DC2 receiver keeps the format of the earlier version; discrete transistors for all but the power amp and R-C low-pass filtering. The filter still allows you to listen to SSB, as there aren't many poles and the cutoff is nearly 900 Hertz —it just removes the ice-pick in the ear often heard in unfiltered DC receivers. You can change the capacitor values for a different cutoff frequency. Applet E performs this function.

Second Preamplifier Stage with TDA7052 final



Above — The 2nd pre-amp and AF power amp. Experimenting with a number of audio stages, I decided on this cascode common emitter / common base amp biased to provide temperature stability, high gain, low distortion + proper termination of the low-pass filter. (The input Return Loss = 19 dB in my 820 ohm bridge set up). Increase the 100 uF filter capacitor on Q3 up to as high as 470 uF if you hear motor boating (low frequency thumping). This stage is prone to feedback since it's directly connected to the power amp. Photograph.

The simple and effective TDA bridged amp has a fixed gain of 40, so this receiver isn't crazy loud, however, it sounds okay. The bypass capacitor on Pin 2 filters hash noise and *can* remove some of the high frequency din from off frequency stations. Experiment to find the best value for your ears; even 0.015 uF might be your preference. I chose a 0.047 uF for my final version.

The 2 uF coupling caps between the power amp and Q3 can be lowered or raised to suit your parts collection. All the audio path coupling or bypass capacitors were "polysomething" types in my bread boards.

This is a base station receiver since the quiescent current draw listening to noise = 37 mA.

For low parts count or beginner's receivers, IC audio power amps make sense; 1 chip and you're done. Consider, for example, the TDA7052 — 2 bridged amplifiers supply reasonable power and headroom in an 8 pin DIP package. A good, but imperfect part. Depending on your goals and abilities, the limitations of the 7052's fixed 40 dB gain and/or the inability to drive grounded loads or insert additional feedback networks may constrain your designs.

Sound Bytes on 40 Meters:

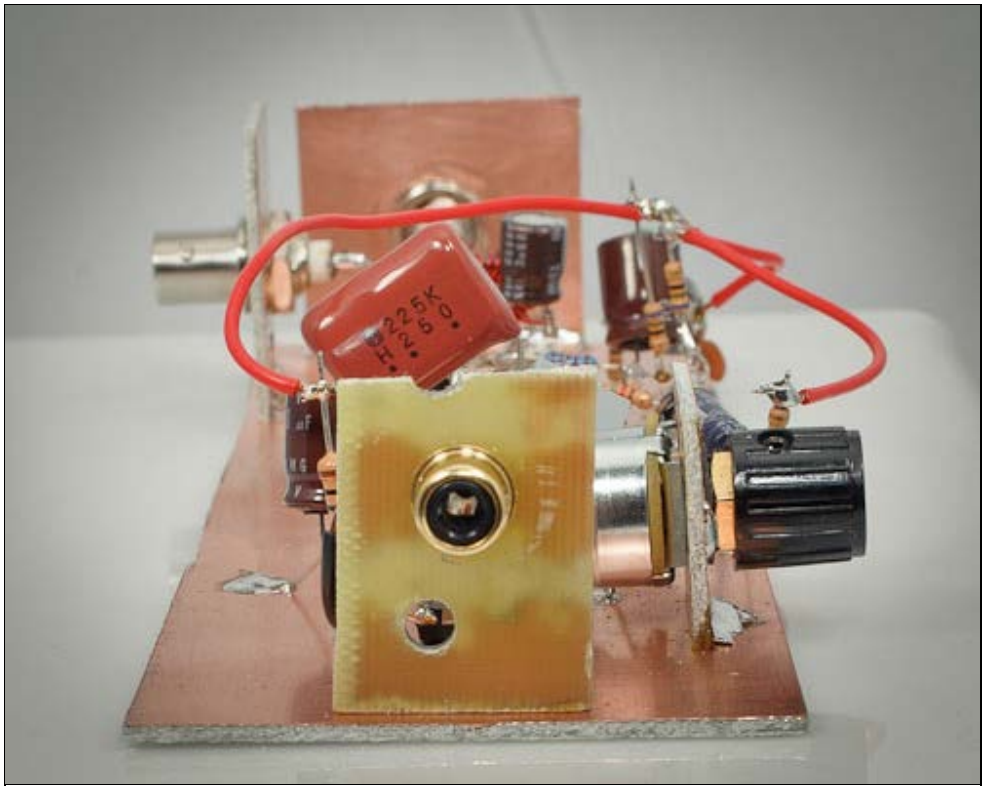
I recorded these sound bytes prior to adding a 0.047 uF bypass capacitor to pin 2 of the 7052 chip and increasing the coupling cap on Q2 from 0.22 to 0.47 uF.

For a control — An ICOM superheterodyne receiver with digital IF filtering set to wide (2.2 KHz) was recorded immediately after recording the Pop DC2 receiver (although I pressed the middle (900 Hz) and narrow (600 Hz) filter selection briefly, but they made the noise worse), The antenna is a 1/4 wave vertical in a city lot with noisy conditions. I don't believe in artificially making my stuff sound better than real, and present warts-and-all audio files. I compressed these files heavily so you'll hear the noise phase shifting a little. Normally with this antenna, a "quiet" QRN level is S9; it doesn't bother me. Icom

Pop DC2 — I was tuning through a pile up to hear how the receiver copes with all the signals (twice as many with a DC receiver!)

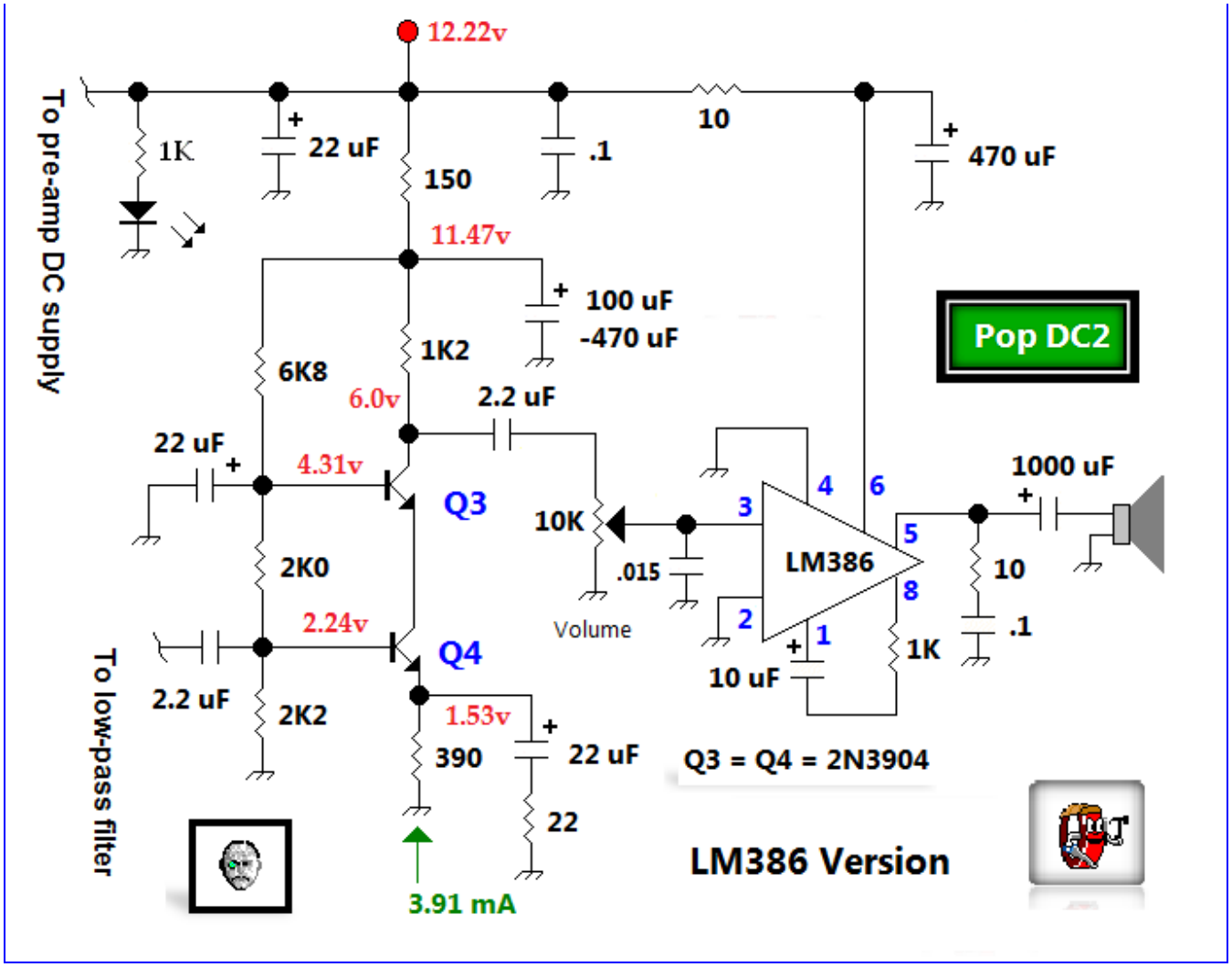
Pop DC2 — More QRN, QSB and pile ups.

SSB - After this, I changed the .22 coupling cap between Q2 and the R-C filter .47 uF to add a little more bottom end.



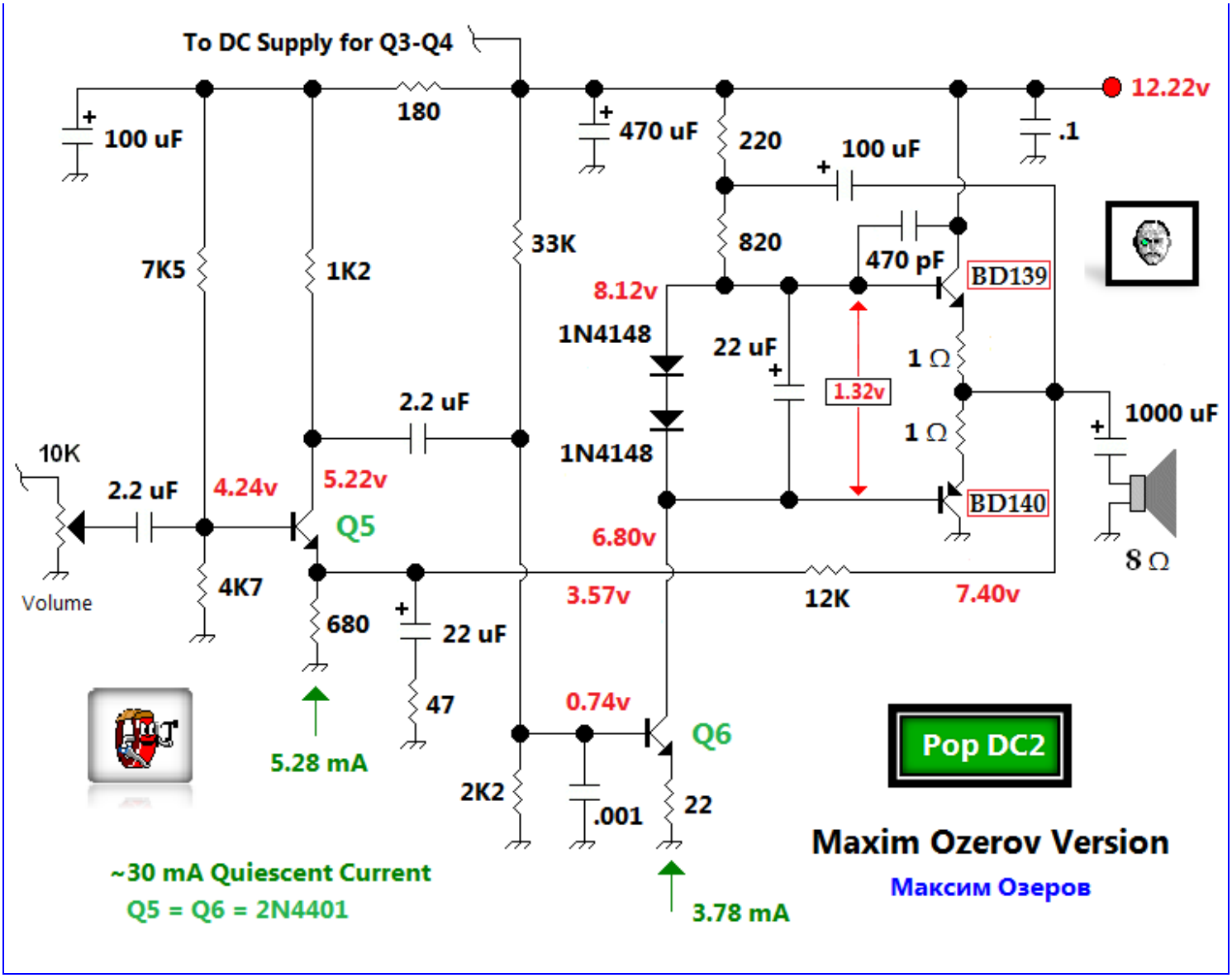
Above — Speaker terminal (an RCA jack isolated from ground on the TDA7052 version). Volume pot at right.

Second Preamplifier Stage with LM386 final



Above — The 2nd preamp and final with an LM386 set for just under a gain of 50. Click for a Canadian SSB sound byte.

Alternate Final Amp Stage that connects to the Q3 volume potentiometer



Above — A reader called Maxim Ozerov requested a discrete semiconductor version of the power amp after I posted the 7052 version on my blog. Placing 2 voltage amps inside the negative feedback loop proved challenging, since I'm no expert and learn on the bench. The gain = ~42 and the maximum pure sine wave power before clipping begins to occur = ~625 mW. This amp is louder and sounds warmer than the 7052 version.

2N3904s work fine for Q5 and Q6, but I found that the 2N4401 had a consistently higher DC beta and this helps ensure the bias and collector resistors shown will provide the widest possible, pure AC signal swing.

This amp replaces the earlier IC power amps (connects to 10K volume pot after Q2, however a 2.2 uF coupling cap is required after the volume potentiometer). If you need more **voltage** gain, increase the value of the 12K negative feedback resistor. Above 75K, the gain will approach 50 and *greatly* increase the possibility of distortion.

My dummy load for development and testing = three 1/2 watt resistors in parallel: 75, 82 and 10 ohms.

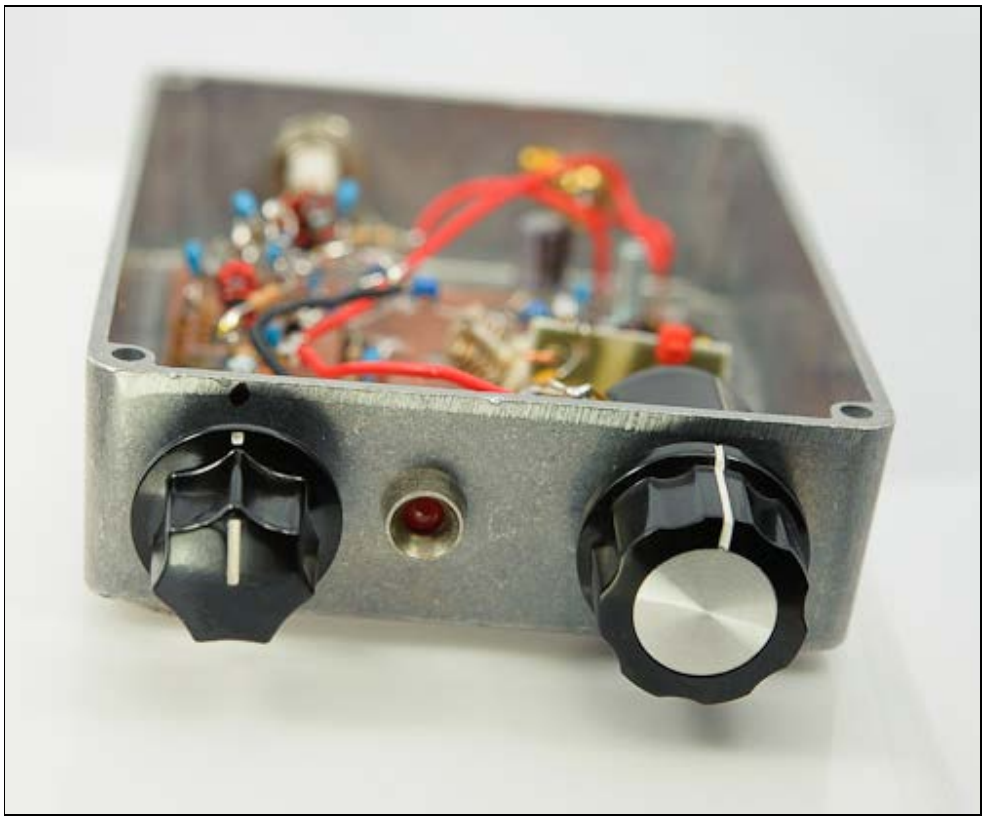
Certainly you can craft better - louder - quieter audio stages with low noise op-amps, however, my readers write that they enjoy building up discrete transistor designs, and for popcorn receivers; I do too.

Sound bytes from November 1, 2011

40 Meters - QRN is lower tonight. Some audio from a lineout tape deck (no tone controls nor equalization). I have only 2 cassette tapes, This one is Russian language from 1983 - the 1 strong accented syllable generates good peaks for AF listening tests. Audio recorded from an 8 ohm, 18 cm (7 inch) speaker mounted in a wooden frame with no back. Speaker choice and cabinets are critical and often overlooked; again we may look to audiophiles for guidance

Click for some 50 ohm AF preamplifier experiments cut from this page.

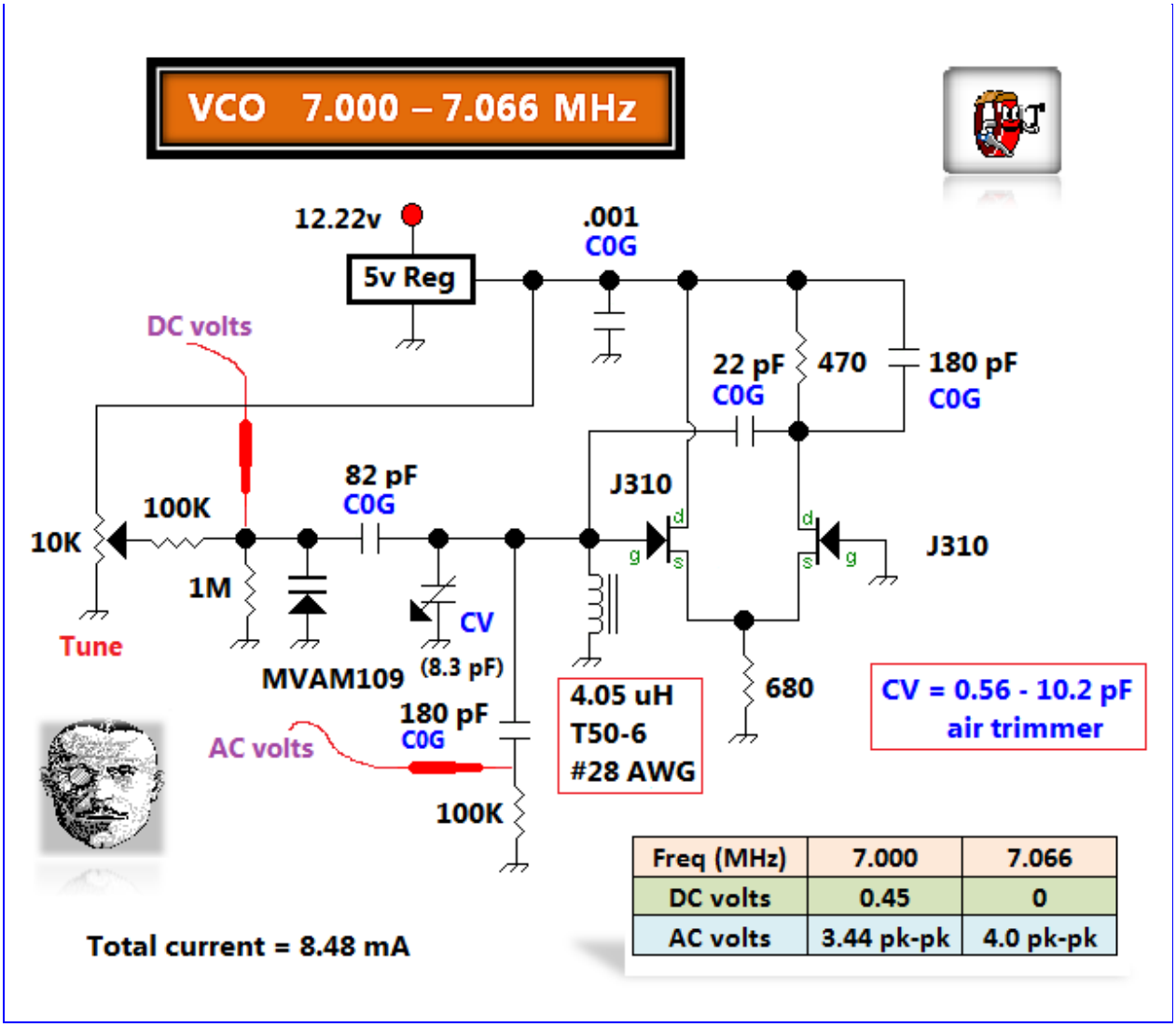
7 MHz VCO Experiments



As RF designers and builders, we rely on signal generators for nearly every experiment. I sought a reliable 7 MHz voltage controlled oscillator and built 1 after some effort. I'll describe and critique a VCO I rapidly designed for a reader and then present a better VCO with some design ideas.

7 MHz VCO Experiments: A rapidly developed Popcorn 7 MHz VCO

A reader needed a 7 MHz VCO in a hurry (3 hours); he only had 1 MVAM109 varactor and wanted to cover the bottom 60 KHz of the 40 Meter Ham band using a linear taper 10K potentiometer for tuning. He planned to use a dual-gate MOSFET cascode buffer (good choice), so I didn't have to bother with a buffer.



Above — The VCO with a 100K resistor as the temporary buffer. He'll use a 100K resistor on G1 of the 2-gate MOSFET buffer. With a Q of 150 at 1 MHz; high noise level and a hyper-abrupt capacitance-versus-voltage curve designed for tuning AM radios, the MVAM109 varactor ranks poorly. The C of my MVAM109 with no reverse DC voltage was 725 pF.

Still, this VCO tuned in a linear fashion, showed a nice sinusoidal output and proved frequency stable. I wanted the AC voltage at the varactor anode at under 1 volt pk-pk (it was 752 mV) to help reduce forward conduction during the positive AC voltage swing. I was bad and ran the tuning DC voltage from 0 to 0.45 volts which greatly increases the potential for forward conduction in a varactor. To mitigate this somewhat, an 82 pF couples the varactor to the tank and drops the AC voltage and reactance seen by the varactor.

In VCOs on the web and print, you'll often see builders connect their varactor to a high Z, and high AC voltage point in the VFO tank; whoa!

At HF, if a varactor is forward biased by the positive half of the AC signal, varactor leakage current and voltage-source loading increases momentarily and lowers Q + broadens tuning. Further, serious harmonic energy and phase noise might be generated as the varactor is biased positive and negative alternately. You can sometimes see distortion in your scope during experiments with extreme AC voltage swings across the varactor. The varactor coupling capacitor should be as low as possible.

Balanced varactor tuning (anode to anode) provides another way to reduce AC signal effects at the cost of reduced maximum capacitance since the 2 varactors are in series. With back-to-back varactors, as the AC signal swings, the varactors are driven into high and low capacitance alternately, but the net capacitance remains constant. Thus applied reverse DC voltage sets the varactor capacitance rather than AC signal amplitude.

The reader for whom I made this impromptu circuit can lower the AC tank voltage by decreasing the VCC or increasing the 680 ohm source resistor after installing the buffer and tweaking things for a 7 dBm output voltage. This topology suffers from an amplitude versus frequency issue — at 7.0 MHz, the output = 3.44 volts pk-pk and at 7.066 MHz the output rises to 4.0 volts pk-pk.

Stuck with an MVAM109 constraint and 3 hours to design/build a VCO, I share this circuit as a raw experiment; not an example of good design because it is not. I took the signal off the gate to derive the best sine wave; this requires a lightly coupled, high impedance buffer with strong reverse isolation to prevent the pulling of the VCO frequency by downstream changes.

A lower L + higher C in the tank, and/or a higher Q varactor could turn this VCO into something reasonable. Popcorn versus high performance? You choose!

7 MHz VCO Experiments: A Suitable 7 MHz VCO

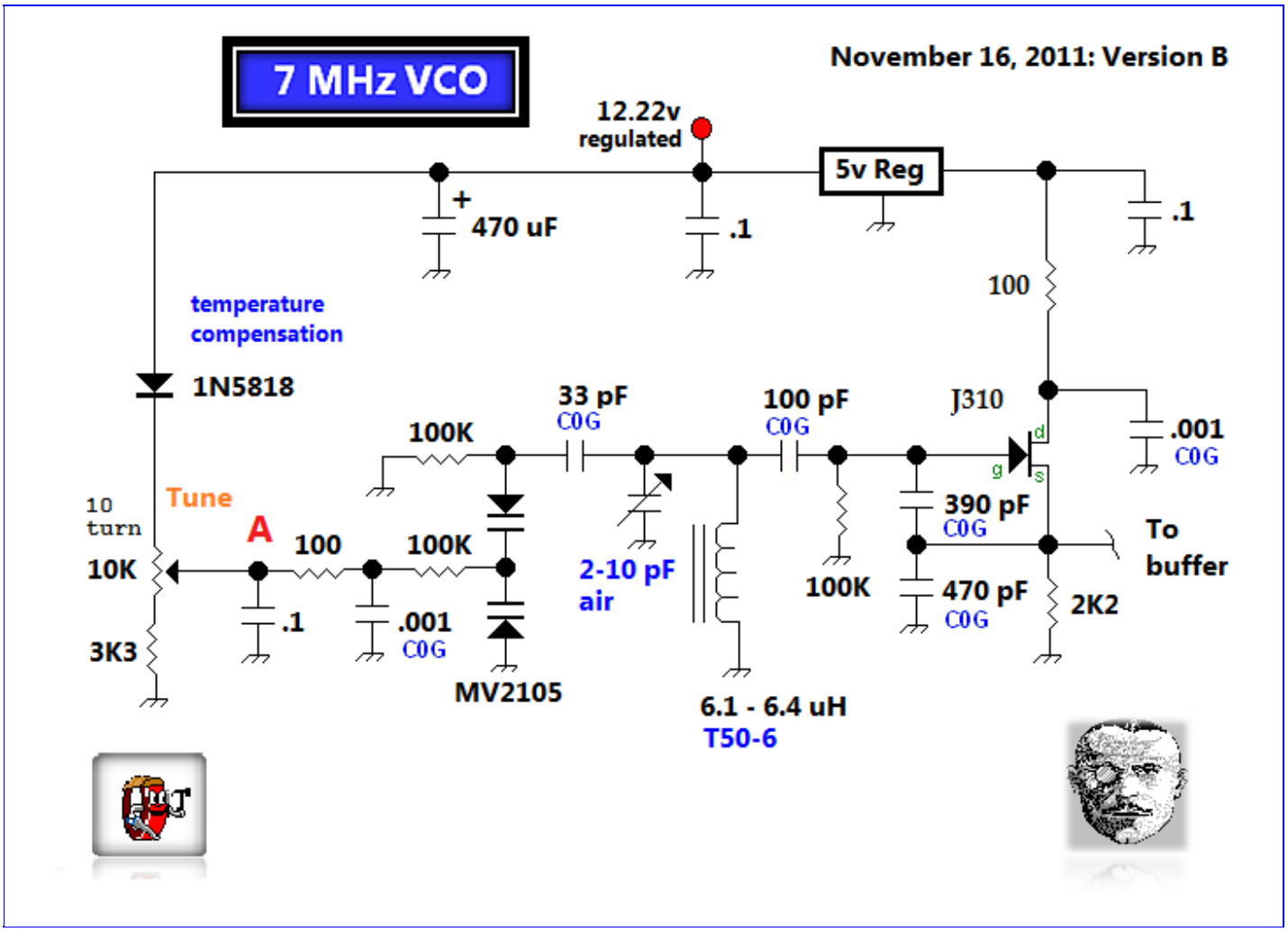
PART 1: Introduction

During my Fall 2011 VCO experiments I studied books including EMRFD and built versions of EMRFD Figures 4.33 and 4.34. Figure 4.33 is a common-base Colpitt's Oscillator using a hyperabrupt varactor. On Q1, the 33 ohm resistor in series with the 0.1 uF cap "de-Q" the 2N3904 to reduce UHF oscillations. Wes also employs current limiting with a 1K5 emitter resistor.

The temperature drift compensation circuit involving a temperature sensitive reference diode + op-amp fascinated me — astute temperature compensation design. I built and tested the whole circuit; the VCO has some amplitude versus frequency and phase noise issues, but it's okay for general use and *great* for varied environments. After tackling Figure 4.33, I built and tested the JFET Colpitts oscillator in Figure 4.34 and share my experiences developing this VCO with an alternate buffer.

These circuits are not cookie-cutter / carbon-copy: they show raw design ideas from the bench.

PART 2: The Voltage Controlled Oscillator



Above — A JFET Colpitts VCO picked after after trying 5 different topologies. This VCO is my version of EMRFD Figure 4.34; originally designed by Wes, W7ZOI.

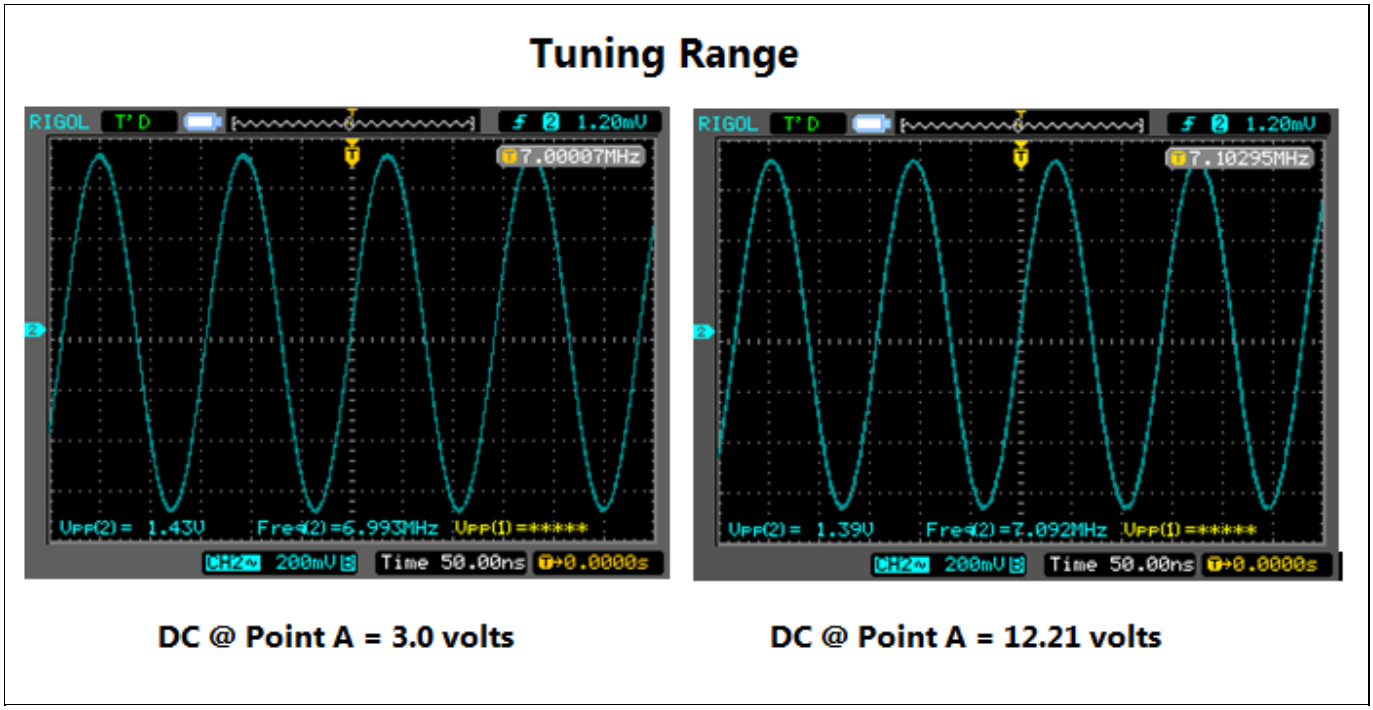
This JFET Colpitts oscillator exhibits a flat output versus frequency, low noise, scales easily to other frequencies and accomodates a wide variety of varactors. For example, you may scale it to other frequencies by changing the L and tweaking the "Colpitt's capacitors" up or down as needed.

I employed a small air variable trimmer capacitor to set the lower band edge and this meant experimenting with the inductor to find 1 that allowed me to set the band edge with such a small trimmer capacitor. I built 2 versions; in 1 the required L= 6.09 uH and in the other, L= 6.4 uH. It would

be much easier to use a trimmer cap with a larger capacitance range as it makes chosing the inductor less exacting.

With the trimmer shown set to half its range, I started with a 6.6 uH coil and remove 1 turn at a time until the output in a counter was close to 7.00 MHz. After permanently fixing the inductor, I tweaked the trimmer cap so the lower band edge was 7.000 MHz with the chassis lid on.

To further drop phase noise, you could reduce the 33 pF coupling cap, add another pair of anti-parallel varactors, run a higher C to L ratio, or perhaps decrease the source resistor to increase the current limiting. Also low resistance, high Q, SMT varactors would help lower phase noise — SOD parts are tiny, but test your hand steadiness and vision.

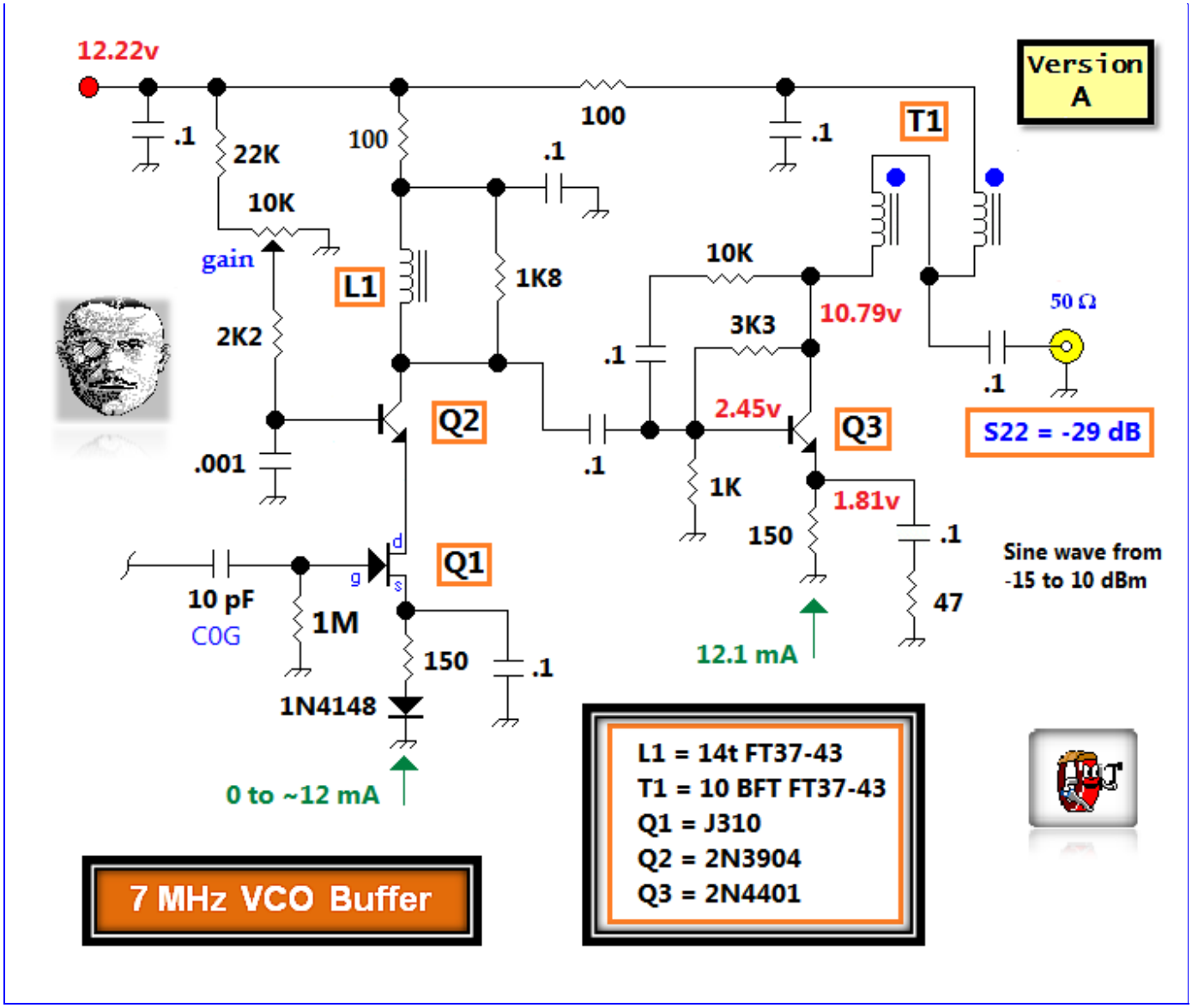


Above — When tuning from the minimum frequency and tuning voltage (7.0 MHz / 3.0 VDC) to the maximum tuning voltage and frequency (7.103 MHz / 12.21 VDC) the signal amplitude only changes 0.04 volts peak-peak.

I kept a minimum of 3.0 VDC on the varactors at the minumum frequency to provide reasonably linear tuning, keep the applied reverse voltage away from 0, and improve temperature stability. All were bench determined and are not factors you can generalize to all VCO circuits. Change the minimum DC voltage on your VCO control by adjusting the resistor on the grounded end of the pot; 3K3 in my case.

Click for a moderate resolution photograph of the VCO and buffer prior to adding the temperature compensation parts.

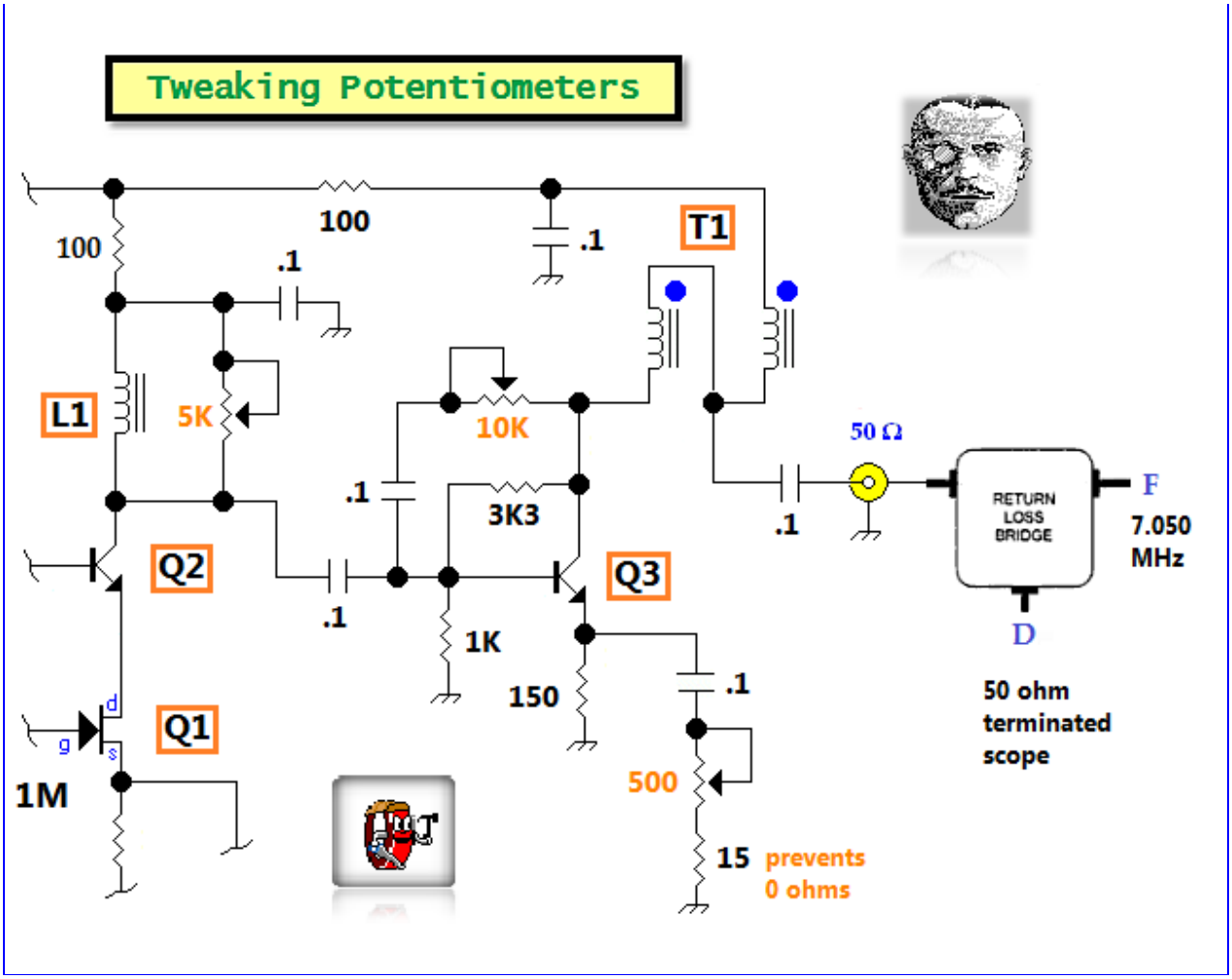
PART 3: The Buffer/Amplifer



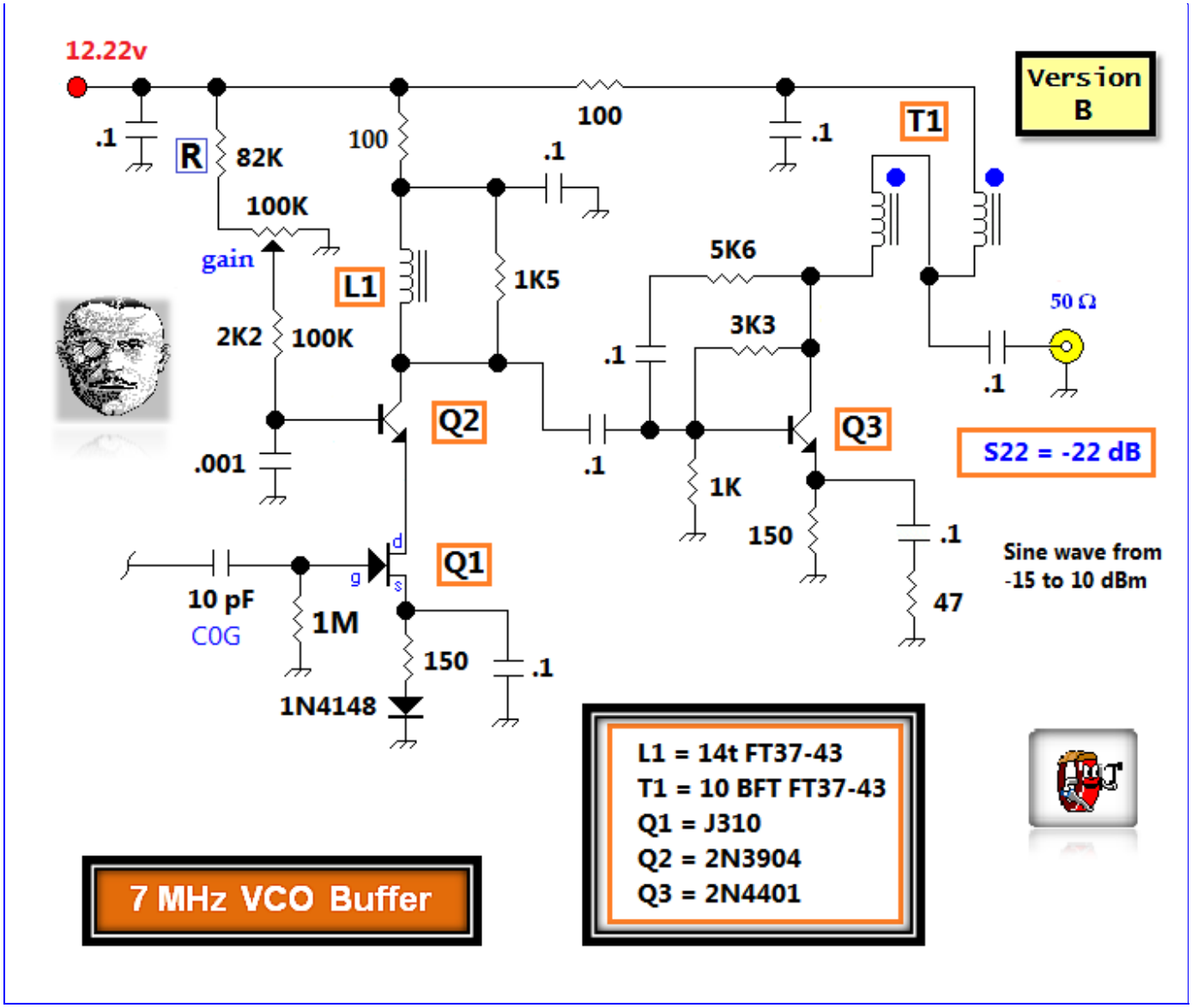
Above — The Q1-Q2 hybrid-cascode amp gives strong reverse isolation (nearly 70 dB) and front panel gain control. You could also employ a dual gate MOSFET or JFET cascode with either fixed bias, front panel control, or a trimmer resistor to adjust the bias on Q2

I enjoyed designing the Q3 final amp amp and matching its input impedance to the output Z of Q2. One way to establish a fixed + known output impedance in order to to get a strong return loss without tuned circuits/networks is to feedback some signal from the collector to the base. The difficulty lies in finding how much negative feedback to apply, while still DC biasing the amplifier for good temperature stability. I set up a crude experiment to determine the Scattering Parameter S22. The goal is to set up a good Q3 output return loss using feedback + matching the Q3 input impedance by tweaking the inductor resistor across L1 and adjusting Q3's emitter degeneration.

The return loss in my first prototype without any attenuator pad = 29 dB; some of this was pure luck.



Above — Q2 and Q3 with 3 variable **orange colored** resistors in-situ and a Return-Loss bridge connected to the output. The potentiometers are tweaked while watching the detected output in an oscillocope. Adjust all the pots for the lowest peak-peak voltage and then carefully remove each pot and measure its resistance with an ohm meter. Replace all 3 pots with the nearest equivalent standard value resistor. Then measure and calculate the return loss (negative of S22). Watch the Q3 emitter resistance — too little R might bring distortion.



Above — A seperate buffer built with 100% different parts that required different AC feedback plus shunt resistor across L1. The parts in this circuit weren't as hot as Version A, and the maximum output voltage was only 1.8 volts pk-pk. In order to get the AC output voltage to just above 2 volts, I had to tweak the resistor labelled **R**.

To keep the heat and current down in the final amp, I decided to keep the maximum clean output to 2 volts peak-peak (= 10 mW = 10 dBm) with an emitter current of ~ 12 mA. If you want higher clean output than 10 dBm, you'll have to run more Q3 emitter current and maybe choose a different BJT, plus apply a heat sink.

When cranked to maximum DC voltage, the Q2 gain pot allows a peak output AC voltage of ~2.2 volts pk-pk into 50 ohms and distortion is evident. At or below 2 volts pk-pk all is well — I'll use this VCO mostly from 0 to 7 dBm.

Since the circuits uses 2 BJTs and a JFET and many 5% tolerance resistors , the Q3 output will vary according to your parts. Tweak the resistor labelled **R** to provide a maximum AC signal just over 2 volts peak-peak into 50 ohms. This translates to around 3.8-5.5 volts DC bias for Q2 with your gain pot cranked fully clockwise.

Return loss variations. You probably noticed the return loss in Version B = 23 dB, while Version A = 29 dB.

Version B originally had the 1K8 shunt resistor across L1 and the 10K + 0.1 uF AC feedback arm just like version A and I measured a return loss of 22 dB. I stuck in 2 tweaking potentiometers (did not bother tweaking the the emitter series feedback element). After pot tweaking, the best return loss I could obtain with 5% tolerance resistors was 23 dB and this probably represents what the average builder will obtain. An S22 of - 22 to -23 dB works fine for the QRP work bench.

If you don't plan to do any potentiometer tweaking, I recommend building circuit A since it has a little more gain due to the slightly higher shunt resistor, and also I built 3 versions of Version A with an S22 of -22 dB or higher.

PART 4: Temperature Compensation

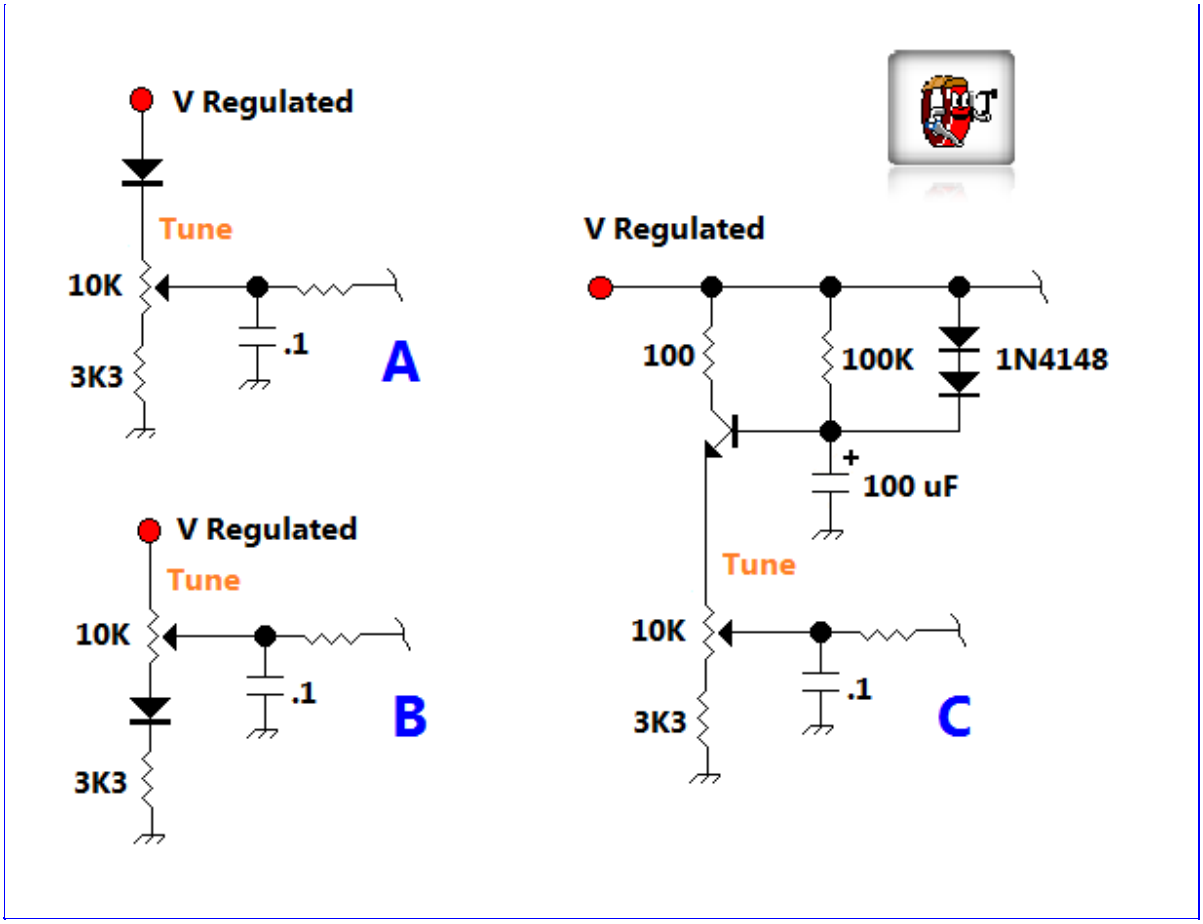
Before temperature compensation, my VCO slowly drifted down in frequency and was unusable.

If you look through the Ham Radio VFO/VCO literature, you will see that many builders use polystyrene caps as the Colpitt's capacitors, and/or in parallel with other NP0/C0G tuning capacitors. Negative temperature compensation caps like an N750, or the polystyrene types temperature compensated the oscillator. Negative temperature co-efficient caps are hard to obtain for many builders; especially in small quantities, however, they are worth their weight in gold.

Diode Compensation

Stabilize your VCO as much as possible with compensating capacitors and by following prudent temperature stability techniques before adding diode compensation. *See the VFO 2011 web page and EMRFD.* Temperature compensation is best performed in a homebrew oven (see EMRFD) and normally takes an incredible amount of time and patience.

Temperature compensating diodes are far from static — a diodes temperature co-efficient is dynamic and may vary with current and also unfortunately, with temperature and even while tuning your VCO !



Above — Simplistic diode temperature compensensation schemes.

The late, great, Doug DeMaw advocated sticking a 2N3904 or 2N2222a (wired as a diode) between the control potentiometer and the varactor decoupling network since the forward biased P-N junction exhibits a negative temperature co-efficient and should stop the decrease in frequency. It can help, however, as you tune and swing the control DC voltage from minimum to maximum the forward bias on the diode increases and the diode temperature coefficient decreases.

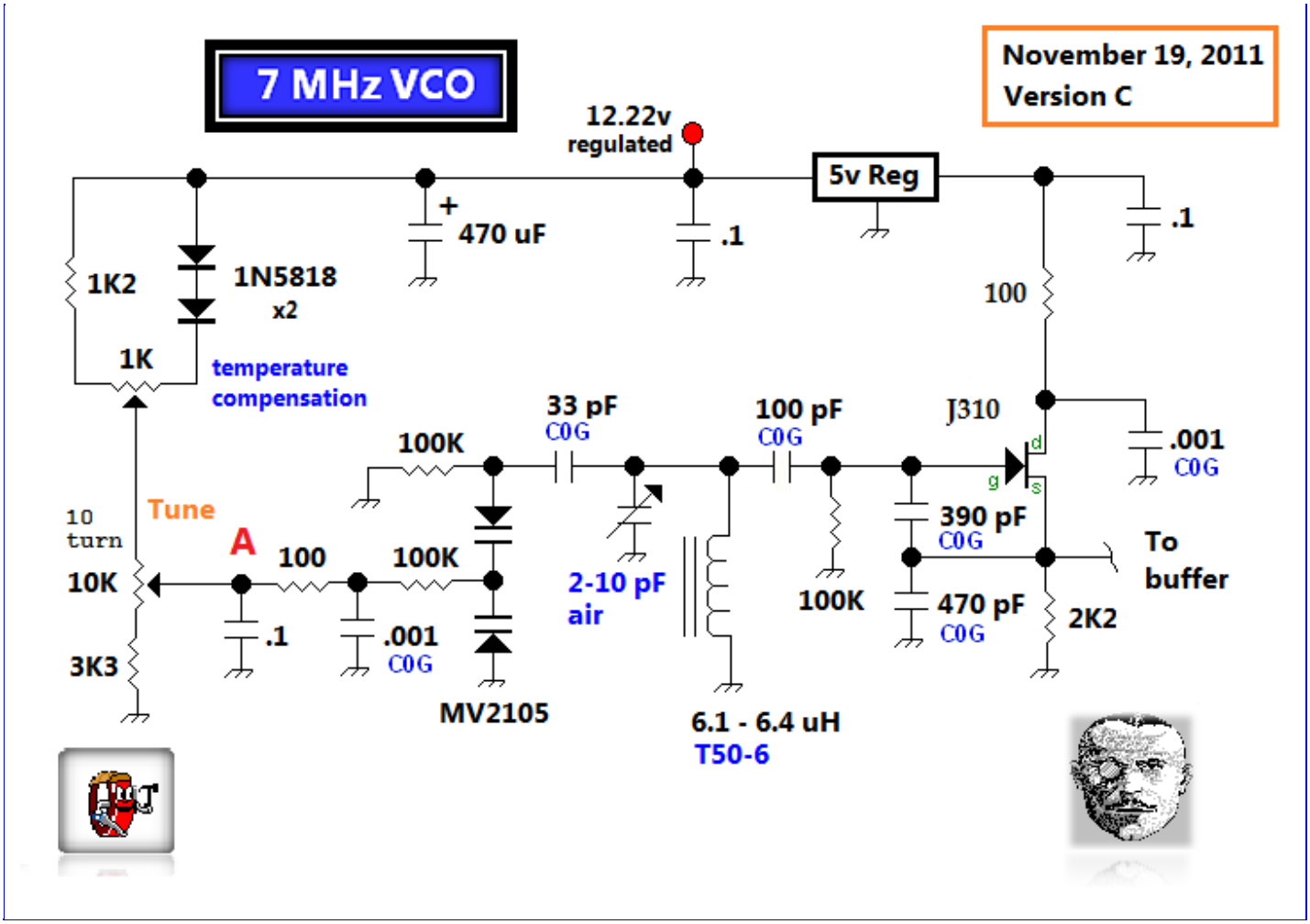
I've never had success using a transistor in this way; the BJT caused the VCO frequency to increase in an erratic manner that varied along with the DC control voltage. When watching drift in a frequency counter set to sample every second or so, a stable design will slowly change frequency in 1 or occasionallly 2 Hertz increments — some people call this "linear drift". if you see your VCO dropping down frequency in 10 - 20 Hertz jumps per second, you'll have a bad time temperature compensating.

I experimented with the above 3 designs that keep a constant current on the diode. Figures A and B work. I tried both and confirmed that a given diode compensated slightly differently when in circuit A or B. This gives you a bit of a tweaking room for your chosen compensation diode. I tried Figure C, but it had too much negative temperature coefficient and sent the VCO drifting upward about 1-2 Hertz each second.

I settled on circuit A and then tried some diodes: the initial best was a grubby old Germanium from my junk box. The best choice turned out to be a Schottky barrier rectifier (1N5818). I connected the VCO to a receiver and could listen to CW QSOs without groaning. My VCO now drifted up in linear, 1 Hertz hops at about 105 Hz per hour. It took a long time to tack solder in and wait 10 or 15 minutes for each diode to stablize before I finally settled on the 1N5818.

The better solution is to choose a suitable diode and vary its current to tweak the temperature compensation. Wes did this in EMRFD Figure 4.33.

Advanced designs may use a reference voltage + a temperature dependent voltage that is applied to op-amps in a proportional way to temperature compensate the DC control voltage. Then, too, some builders ovenize their VCO container to maintain a very stable environmental temperature.



Above — A simple and elegant diode compensation scheme proposed by Ken Kuhn. Basically, it lets you tweak the degree of compensation to what is really needed rather than accept what you get from a diode. Adding more diodes will increase the effect — but the 1.2 K resistor should be increased accordingly to roughly match the overall voltage drop of the diodes. Hopefully there is a point on the 1K potentiometer where temperature compensation can be very good at a tuning point of interest. The diodes should be located to thermally match the rest of the oscillator circuitry.

Set the band edge *after* finding the sweet spot on the 1K potentiometer since it will affect the tuning frequency. This experimental circuit cannot be casually copied and it took a while to converge to the desired operating point on the 1K potentiometer. Generally you start with the 1K pot towards the 1.2 K resistor and then adjust for the best stability after warm-up. Then repeat and adjust as necessary over time.

Temperature compensatiing an oscillator like this is a challenge as all parts have some temperature drift and it takes a lot of measurements (and often, some dumb luck) to determine the overall compensation curve that is needed.

My 1 hour drift up in frequency is now ~ 60 Hertz per hour at various tuning frequencies across the tuning range. I stuck with the 1N5818 diode, and probably should have tried other diodes and also changed the 1K2 resistor to observe any effects, however, I have spent an inordinate amount of time on this circuit and leave it to others, or future bench work to improve. *See QRP — Posdata below.*

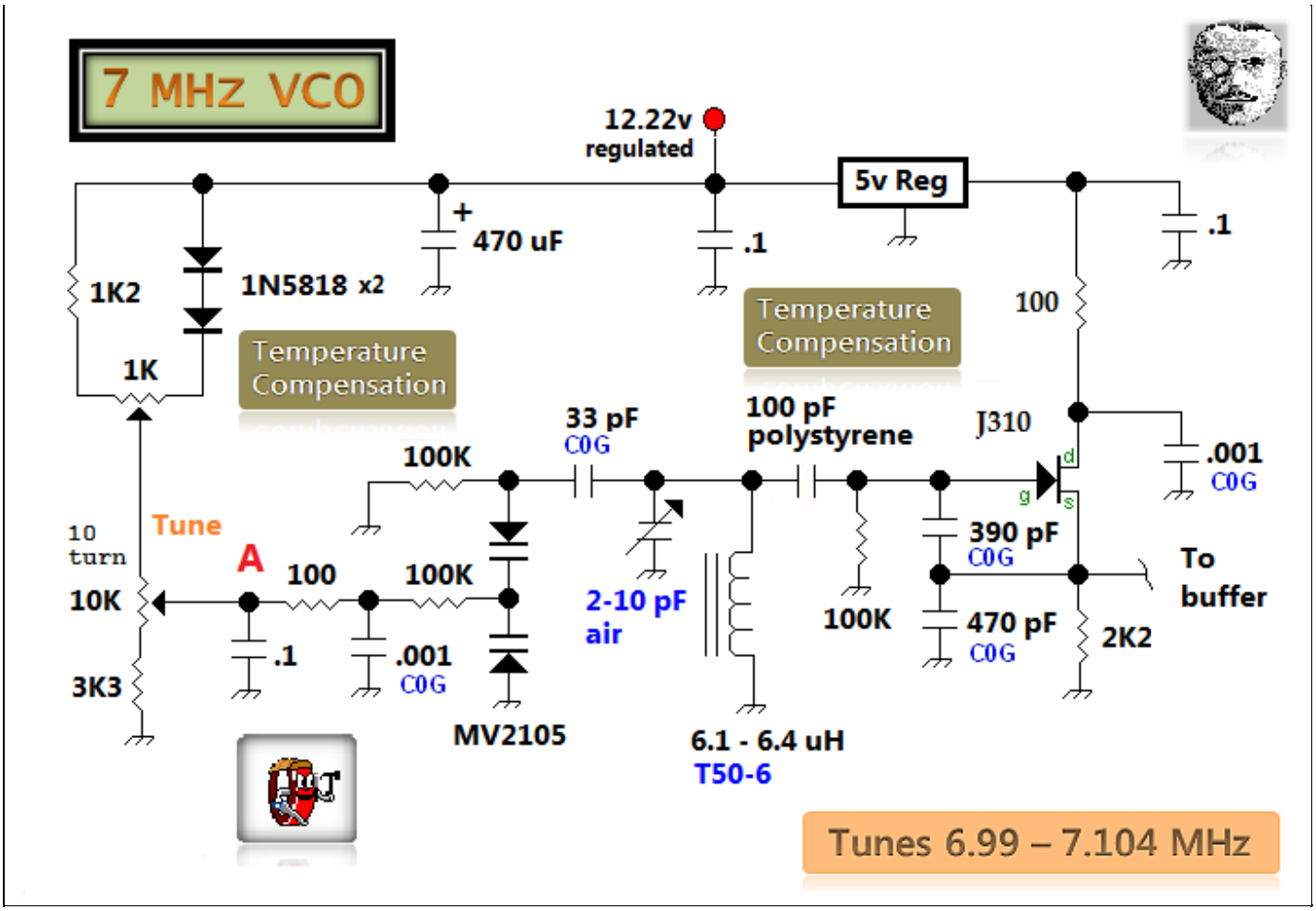
Above — Version C of the VCO. When I built the first versions, I drilled a hole to accomodate a third potentiometer, but filled it with a LED holder that was temperature sealed with epoxy glue. The 1K pot *jus*t fit into my chassis.

QRP— Posdata for January 2012

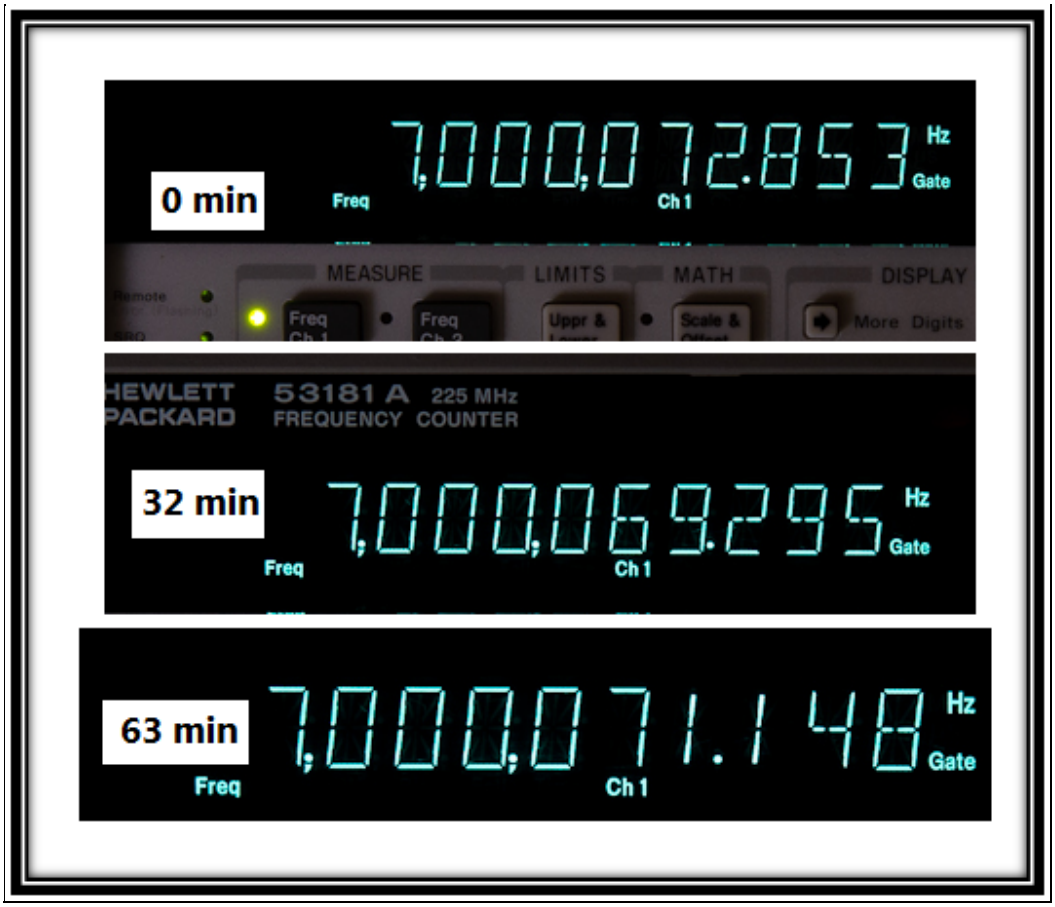
In late 2011, I shopped on eBay to build up a small quantity of 10 - 270 pF polystyrene, plus some 56 pF N750 ceramic temperature compensation capacitors. After 2 simple, but time-consuming experiments, I temperature stablized my 7 MHz VCO frequency drift to under 10 Hz per hour in the relatively constant temperature of our basement.

I didn't feel like re-doing the whole resonator circuit and thus focused on the tank to FET coupling capacitor. Placing a N750 capacitor in parallel with a fixed NP0/C0G cap to make 100 pF resulted in over-compensation and no amount of tweaking on the "adjustable diode" circuit worked. A few hours late, I swapped in a 100 pF polystyrene capacitor and after further hours of waiting and tweaking, I nailed the frequency stability sought.

An oven provides the best way to temperature compensate, however, whether you choose the oven method, or the bench method like I did, great patience is required to see if a change to your TC circuitry works or not.



Above — Final version of the 7 MHz VCO. I changed the 100 pF capacitor coupling the JFET to the resonator circuitry from C0G to polystyrene and slowly tweaked the 1K temperature compensation pot to find the point of convergence.



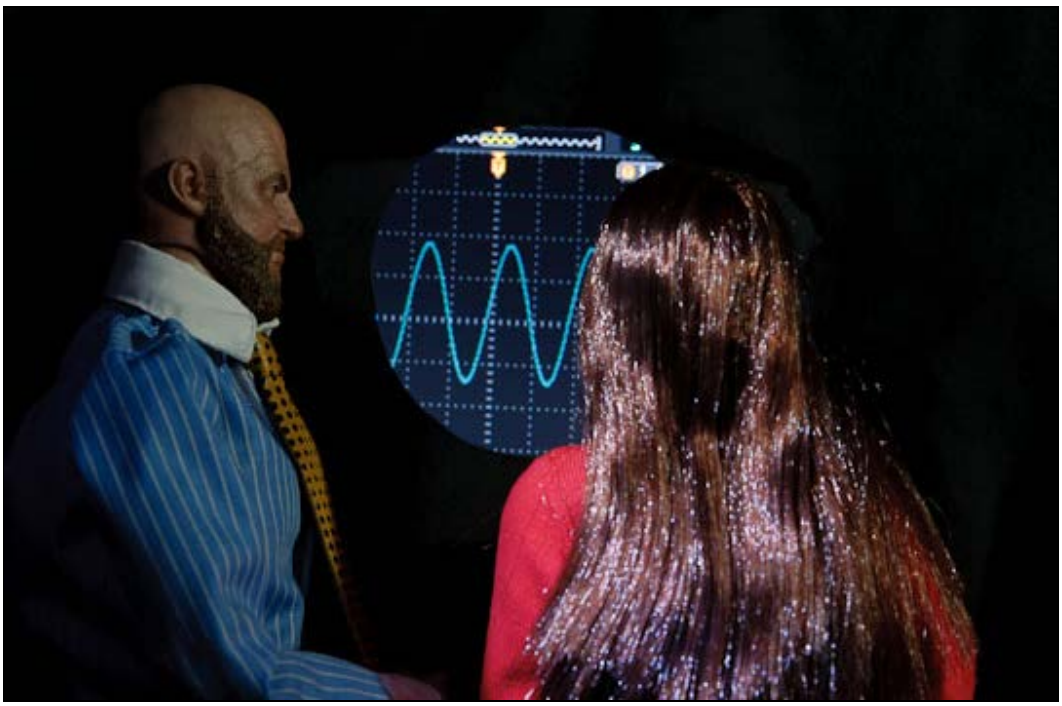
Above — The ~1 hour drift after a 30 minute warmup period for the 7 MHz VCO. Love this.

Above — My new temperature compensation capacitor parts drawer. I'll keep an eye out for further bargain temperature compensation parts on eBay and at Ham Radio festivals.

Fearless Leader and Hero Храбрый вождь и герой

Above — Professor Vasily Ivanenko ( ), fearless leader (ТЫ МОЙ ГЕРОЙ)

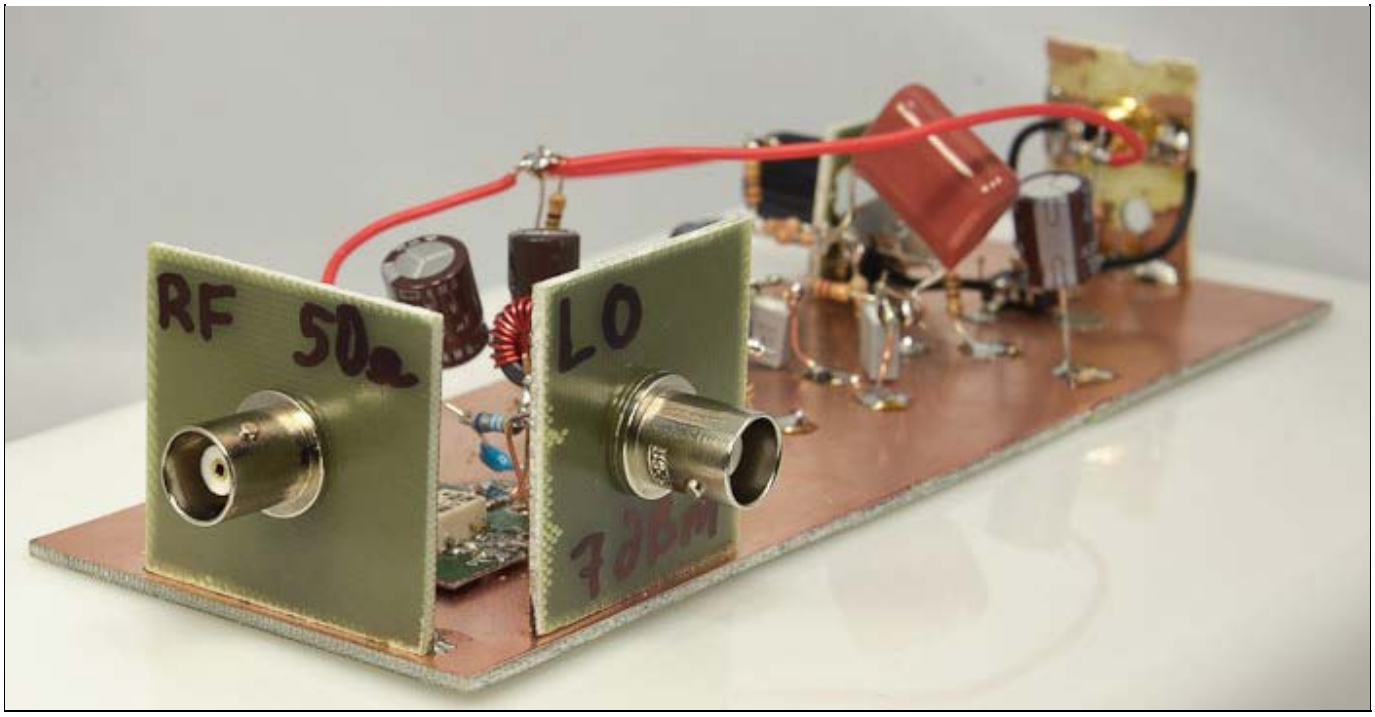
He's my hero because he's humble, fallible, well-intentioned and moral. Professor Ivanenko lives for learning — fame is filler — hollow and distracting. His current *ego* lags his voltage by 90 degrees. Is he part inductor / part human?



Miscellaneous Photos and circuits

Above — 3 types of adapters. A BNC male to SMA female, a BNC male to PL-259, and a BNC female to S0-239 allow the RF modules to be connected to a variety of equipment.

Click


RF 50Ω
LO
7dBm

Double Tuned Receiver Band-pass Filter Design Center

This web page is for builders who own EMRFD. Assisted by 4 of the Ladpac programs from the EMRFD compact disk, and the information presented in EMRFD Chapter 3, I share some experiments building popcorn receiver band-pass filters. Prior to diving into this material, please read the help file ***Ladpac2008 Manual.pdf*** and a file on the EMRFD compact disk called The Double Tuned Circuit: An Experimenter's tutorial by Wes, W7ZOI.

Preface

Derived from experiments, my web content reflects the efforts of a lay-person, hobby-level designer — I make mistakes. I say this not to make excuses or avoid accountability, but to share the truth. My hope is that my experiments inform yours and we all improve over time. I correct reported mistakes and rely on your eyes to see them.

Arduous and requiring good math skills, filter design is out of reach for many builders. Software changes this and learning to apply computer programs in real-world situations is part of our hobby. This web page shares some bench experiences, plus my thoughts about using some programs written by Wes, W7ZOI. I present suggestions and examples based more on empiricism and from reading about band-pass filter design than scientific methodology.

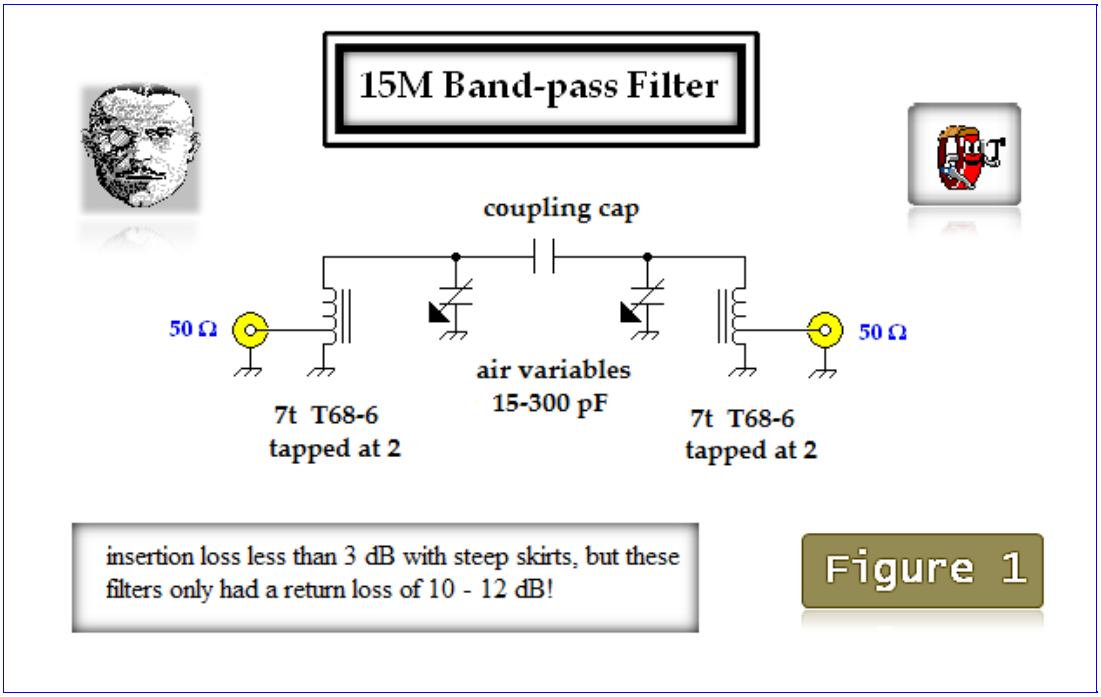
From email regarding my VFO and RF Workbench pages, I have become aware that I've lead many builders to think that a perfect sine wave and a high return loss are "must have" bench outcomes. This is false. A clean sine wave proves useful for accurate measurement, but is not a *de rigueur* bench outcome. A desire for high return loss reflects my own personal *obsession*; in simple QRP rigs, this may represent folly. Please don't overestimate the importance of return loss from my bias; decide for yourself.

Part 1: Experiments with 2 coupled L-C tanks.

Goal: A 15 Meter band band-pass filter with an insertion loss < 4 dB and a return loss of >= 20 dB.

Software: Ladbuild08 and GPLA08.

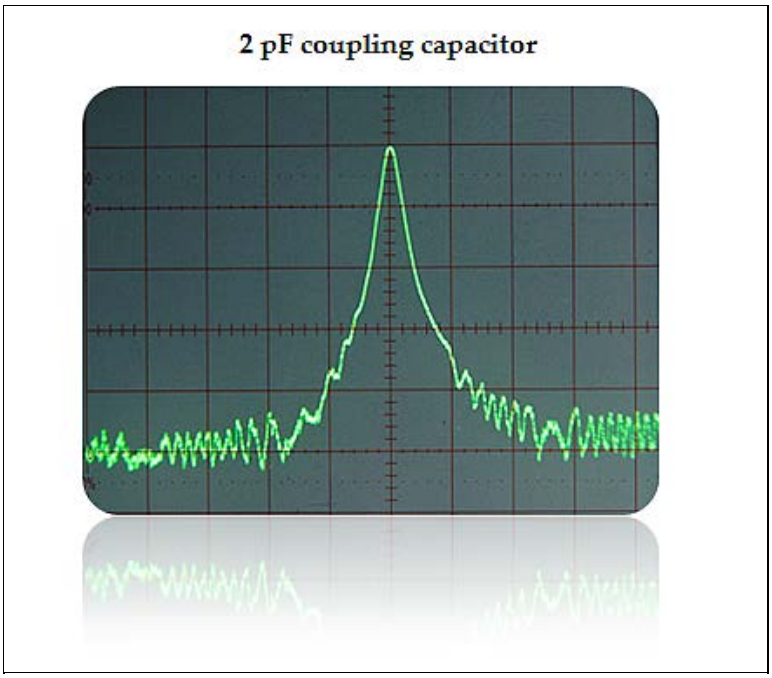
The simplest band-pass filter is an L-C tank. To get a decent stop band we generally couple 2 or 3 tanks together with series capacitor(s). Other filter topologies were ignored. In Part 1, I just connected up a couple of tanks on the bench without the use of software. Some attempts at impedance matching via transformer links were also trialed.



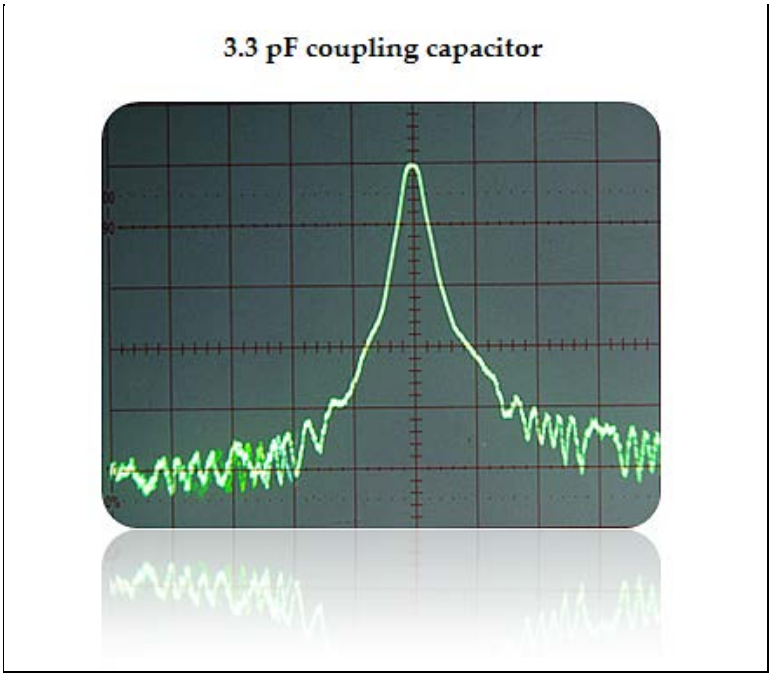
Above — I built a 5 component filter for the base experiment. Inductors = 7 turns of #22 AWG on a T68-6; tapped at 2 turns from ground. The inductors turns were expanded or compressed until L= 300 nH. Tuning capacitors = large 15 to 300 pF air variable capacitors. Coupling capacitors trialed = 2 pF, 3.3 pF, 5 pF, and 7.5 pF.

After soldering in a coupling capacitor, each tank (also called resonator) was tuned to resonance by looking at the peak-peak output voltage in a 50 ohm terminated oscilloscope. After tuning, I measured insertion and return loss and then swept each filter with a tracking generator + spectrum analyzer. On the bench I determined that the greatest return loss occurred with 2 transformer taps from ground; the result — a dismal 10-12 dB.

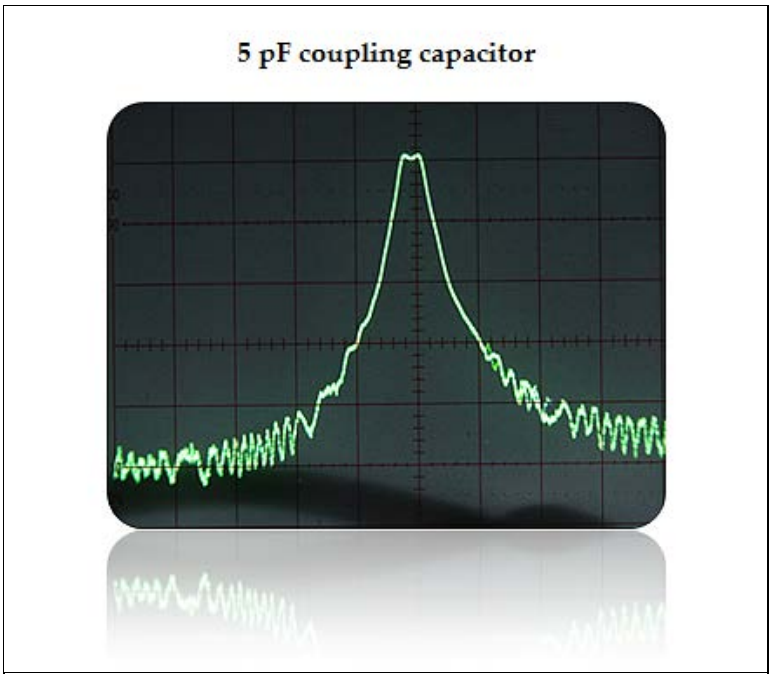
What effect does changing the coupling capacitor have?



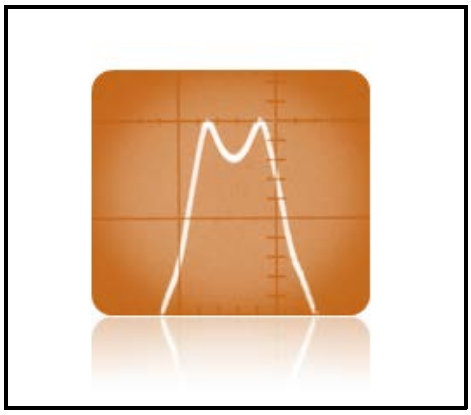
Above — A spectrum analyzer + tracking generator sweep of the filter response with a 2 pF coupling capacitor between the inductors. Graticules = 2 MHz per horizontal division and 10 dB per vertical division. Click on this zoom to better see the - 3 dB bandwidth. The sweep revealed a sharp peak response with steep skirts and a 3 dB down BW of ~220 KHz or so. Some of the noise arose from the big air variable caps connected to each tank with short hook up wires, plus no shielding.



Above — The SA + TG sweep with a 3.3 pF coupling capacitor. The peak isn't as sharp, but still looks good. As shown, increasing the coupling capacitor value increases the 3 dB filter bandwidth with all other components equal.



Above — With a 5 pF coupling capacitor, a double humped response appeared. The bandwidth further increases.

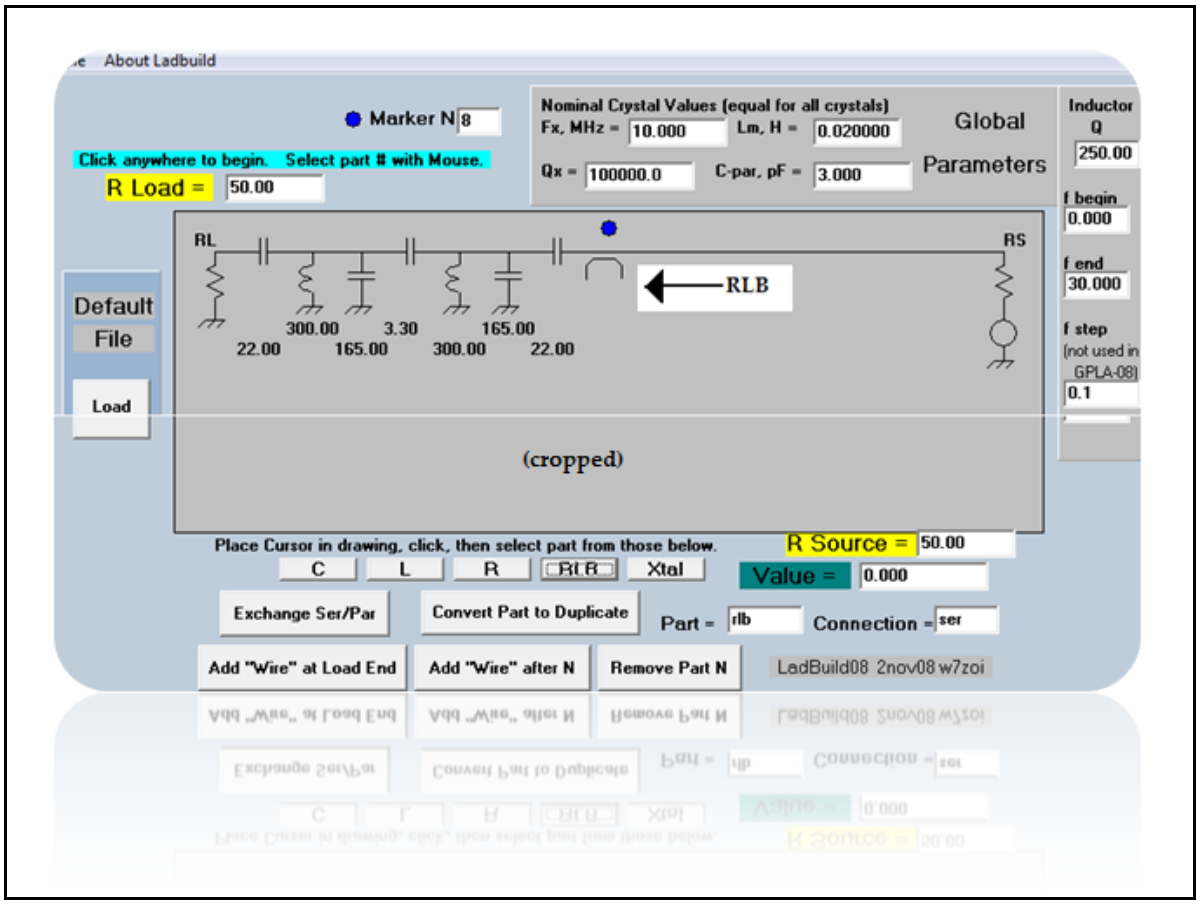


Above — A zoom of the double humped filter response employing a 7.5 pF coupling capacitor. Imagine the difficulty tuning this band-pass filter in a receiver by listening to band noise. Tuning in either peak skews the filter bandwidth. Additionally, the 3dB bandwidth now = ~ 1.6 MHz — Not a good filter!

Optimizing Return Loss

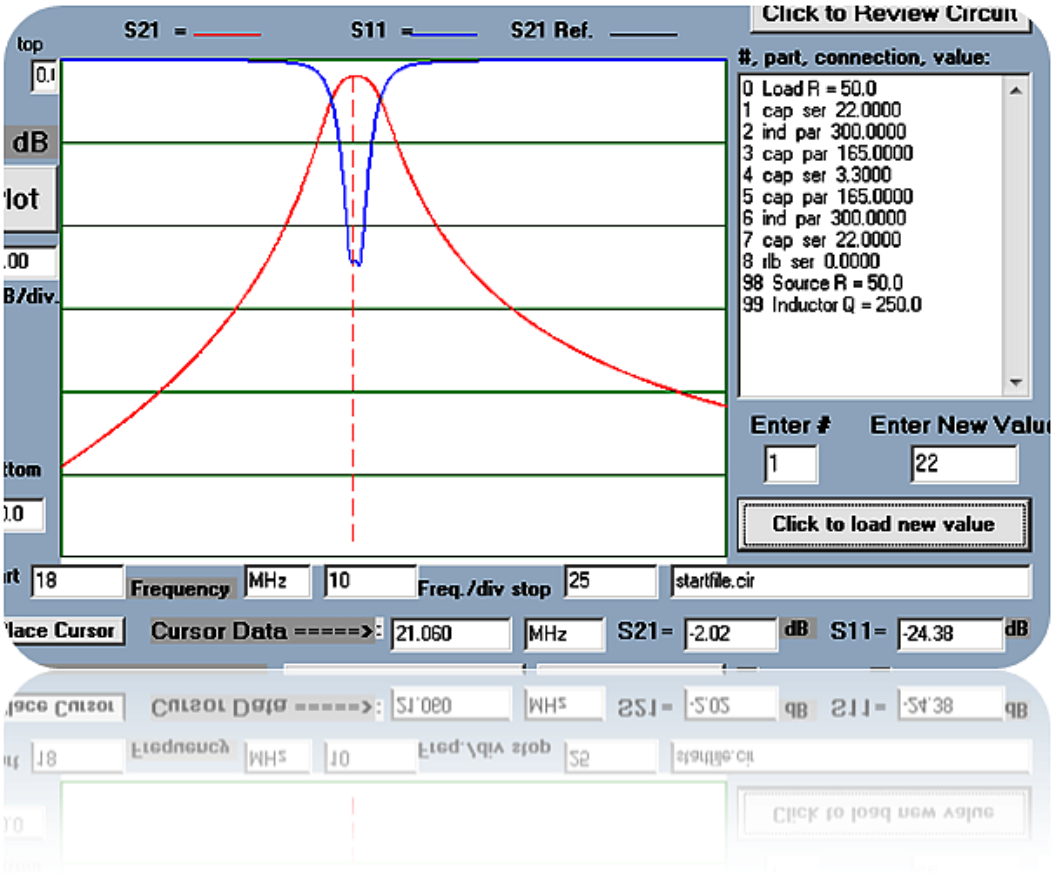
Despite trying, I could not obtain a better return loss than 10 -12 dB by changing the tap point on the 7 turn inductors. In part this was due to limited potential autotransformer ratios on a 7 turn coil. I emailed Wes, W7ZOI and he sent this file. I learned that adding a series capacitor to each end will tune the filter to 50 ohms impedance. What capacitor value should we use?

The answer can be found purely experimentally, or with Ladbuild08 to make a digital file of your filter and GPLA08 to analyze it.

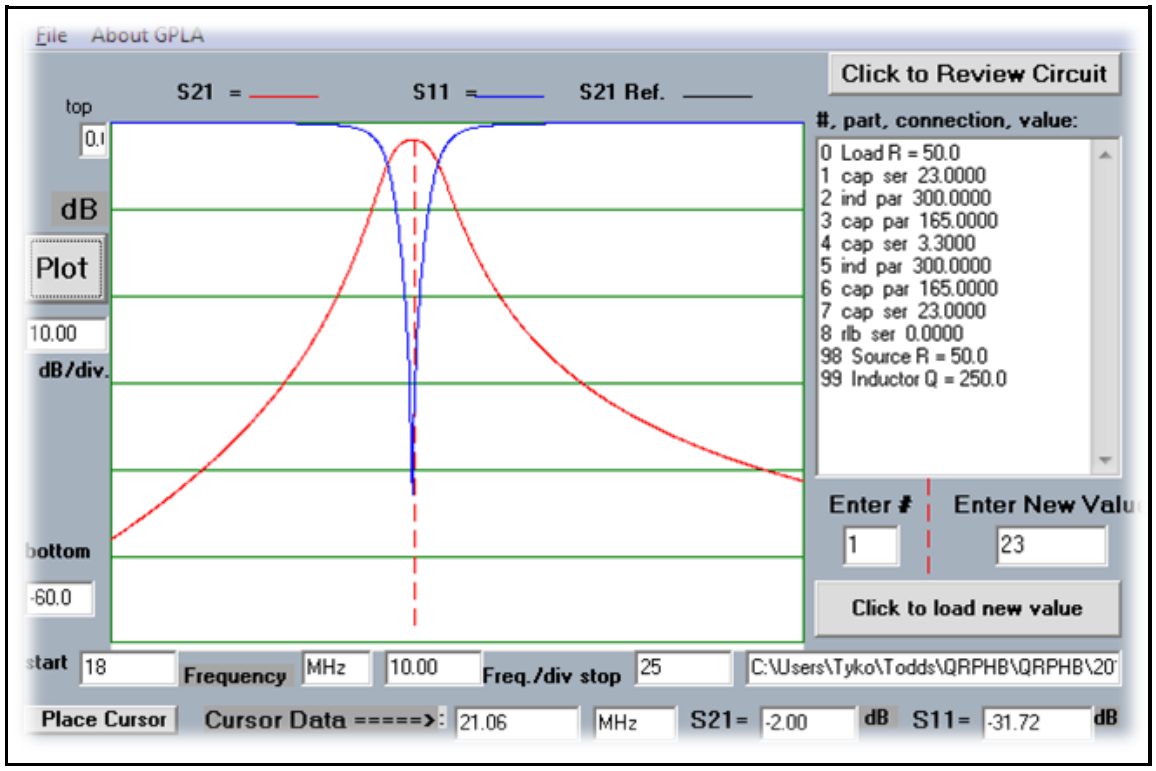


Above — I "built up" my Figure 1 filter in Ladbuild08 with a 3.3 pF coupling capacitor. Initially I guessed at the values for the series end capacitors and knew my tuning capacitor were ~ 165 pF because I removed and measured them from the peaked filter from Figure 1 and added a few pF for stray capacitance. Any of these values can be changed in GPLA, so educated guessing is okay.

For size 50 to 68 toroidal inductors, many builders choose a Qu value from 200 - 250 with # 6 material. Qu affects insertion loss and to some extent, return loss. Click for a tutorial from Wes', W7ZOI site and consult EMRFD for more information.
In order for GPLA08 to display an S11 plot (return loss), a return loss bridge (RLB) must be added as shown. Also check the **Plot S11** check box in GPLA08.

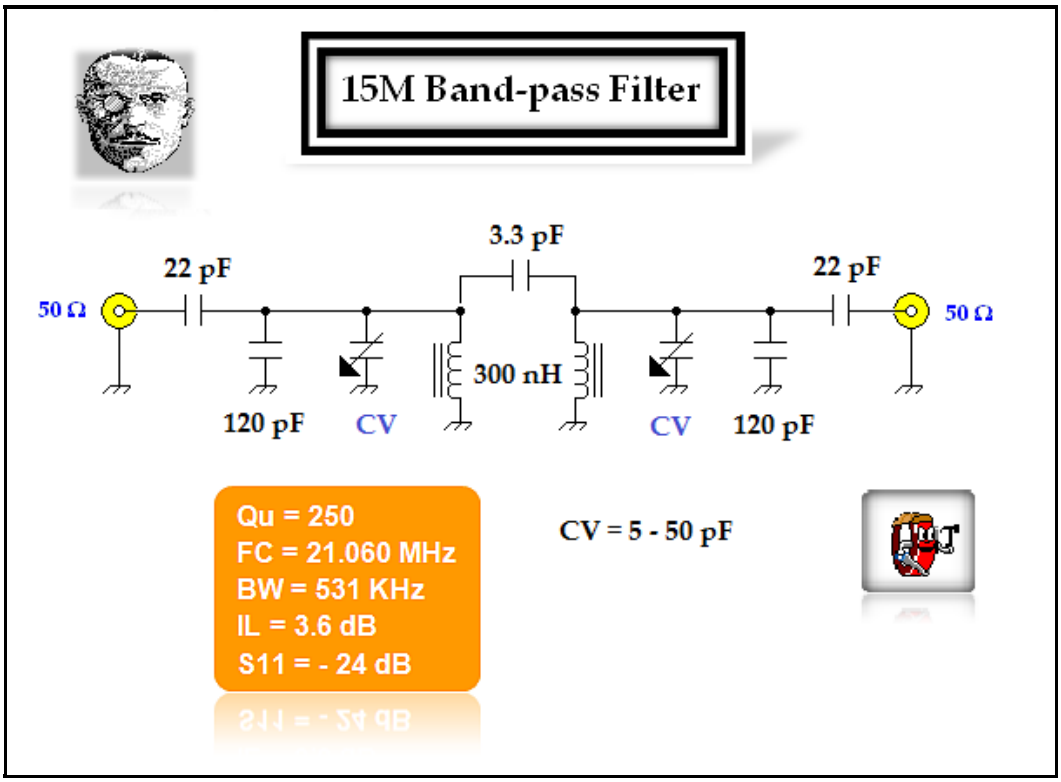


Above — The GPLA 08 filter simulation of the filter "built" with Ladbuild08 above.



Above — In this filter simulation, I tweaked the end capacitors (parts #1 and #7) from 22 to 23 pF and watched the return loss (S11) improve by 7.34 dB — if wanted, you can optimize the end capacitor values to improve the match into 50 ohms. To re-establish the center frequency, slight retuning of parallel capacitors #3 and 6 is required when changing the series end capacitors; although I specifically didn't change them for this example.

Increasing the 2 end capacitors to increase S11 renders an option only; you don't have to go for the best S11 in your filters. Increasing the series end capacitors to bump up return loss tends to increase the 3 dB bandwidth and reduce insertion loss.



Above — I built and measured the filter with 22 pF end capacitors since these are common, standard values. In another experiment, a 1 pF cap was soldered in parallel with each series capacitor and the return loss increased by about 4 dB. Click for a bench photo of an alternate version of the above filter. Clearly, GPLA08 simulation furnishes us popcorn builders with a starting point to make top-notch band-pass filters.

Click for another simulation of a filter employing a 2 pF coupling capacitor, with the end capacitors tweaked for the best S11. *S-11 is just the negative of the return loss.* I would certainly use this filter in the front end of a popcorn direct conversion receiver.

An easier way to design your band-pass filters involves using DTC08 to design a raw filter and GPLA08 to substitute in standard value capacitors and tweak your filter. That's part 2. The material presented in this section supports the discussion in Part 2 and 3.

Part 2: Band-pass Filter Design using DTC08

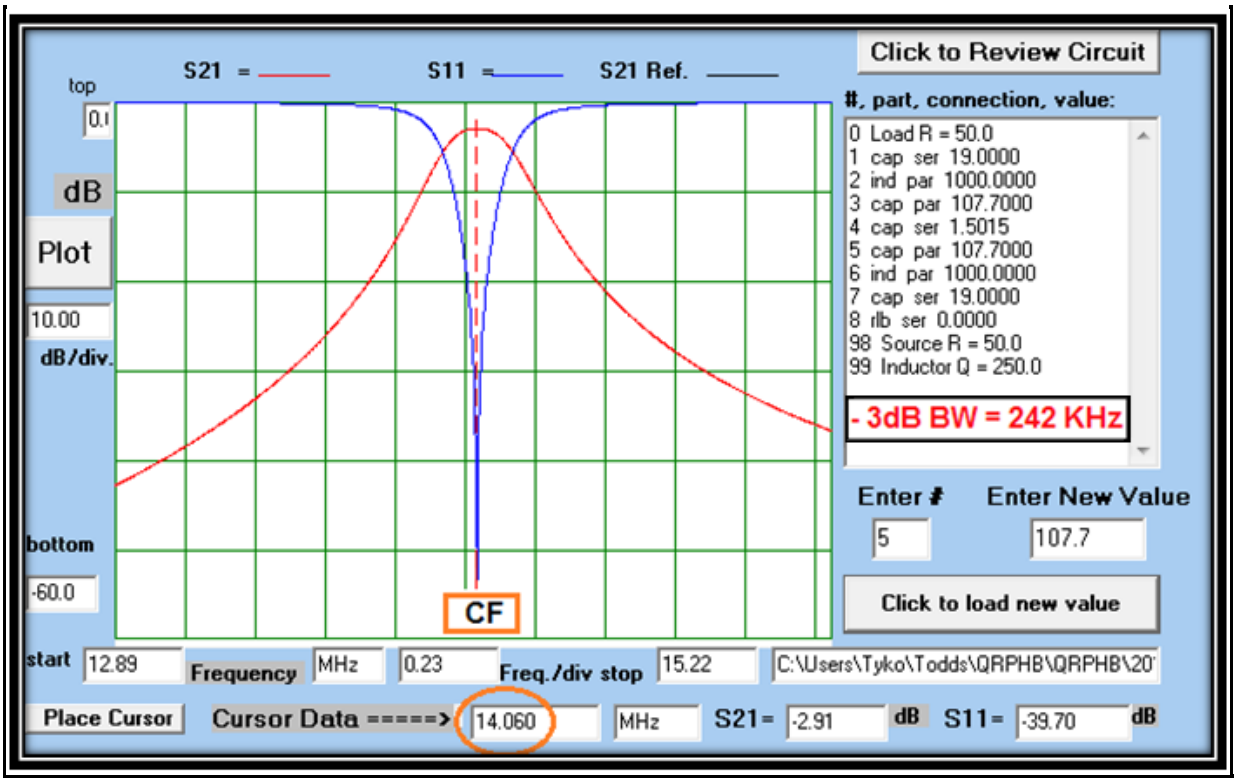
Prior to using these Ladpac programs, some numbered design points and a preamble follow.

More than anything else, our parts collection dictates what filter parameters we choose and end up with. For example, if you want filters with a low bandwidth such as 150 KHz and under, you'll require inductors and capacitors that provide really high Q, or you might suffer from punishing insertion loss.

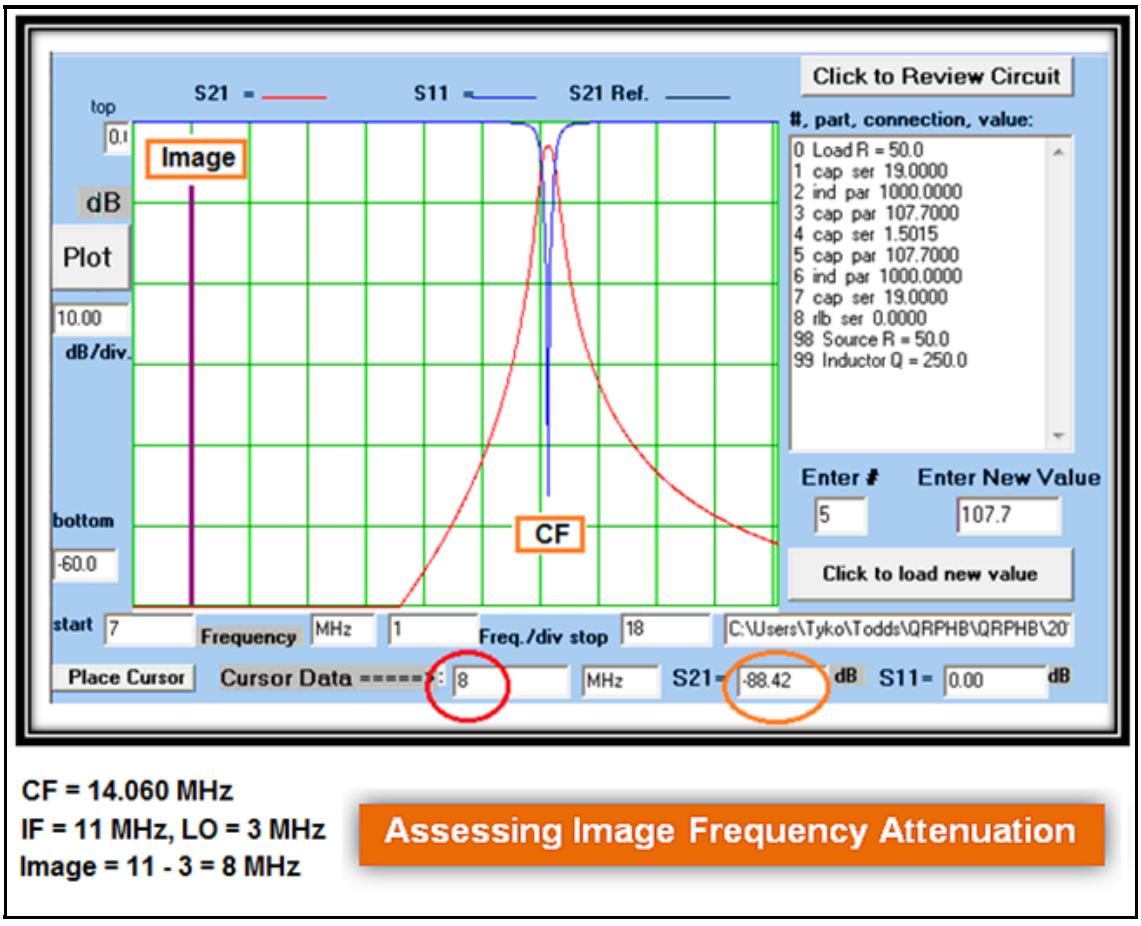
The following are general starting points only — *your* needs, parts and abilities drive *your* filter design. Example variances include: if a low noise amplifier follows a filter, a higher insertion loss might be okay; a high return loss is not always required for a low noise figure; especially in popcorn receivers. Also, it's a viable choice to trade off insertion loss for steep skirts in some filters.

1. A reasonable 3 dB bandwidth = 100 to 500 KHz, but this depends on the purpose of the filter.
Numerous considerations challenge us. Will this be a whole band (CW + SSB) filter, or a CW only filter? As a CW op who uses simple equipment, I tend to design moderate bandwidth (200-300 KHz) CW-only filters. If you need CW + SSB, then a bandwidth of 350 KHz or greater might suit you. It's really up to you.

Other factors affecting bandwidth choice include whether the filter drives a superheterodyne or a direct conversion receiver. In superheterodyne receivers, your intermediate frequency informs your filter bandwidth choice. Consider the following 2 diagrams:



Above — Using DTC08 and GPLA08, I designed an example filter for the front end of a 14 MHz superheterodyne receiver with an 11 MHz IF. BW = 242 KHz. CF = 14.060 MHz; a frequency some QRP operators favor.



Above — Assessing filter attenuation at the image frequency using GPLA08. *To keep the arithmetic simple, I employed a frequency of 14 MHz for the image frequency calculation.* As shown, the simulated attenuation of my 8 MHz image frequency is 88.42 dB. Since I personally target an image frequency suppression of 60-70 dB; at 88 dB, if I wanted, I could increase the bandwidth of this filter for broader coverage and reduced insertion loss.

How much image frequency rejection is needed for superheterodyne receivers? I'm uncertain, for I have seen competent authors choose between 50 and 100 dB. I feel a good target = 60 - 70 dB, and 50 dB is the bare minimum. To realize image attenuation above 50 dB, shielding is usually required.

Three or more L-C tank band-pass filters may be required when your image frequency is close to the IF frequency. Choose both your intermediate frequency and your bandwidth wisely.

2. After selecting your bandwidth, *tweak the inductance* and only if necessary, make minor adjustments to your set 3 dB bandwidth to give standard, or near-standard value coupling capacitors that you *own.* Obviously, you can place fixed capacitors in series or parallel, or even couple your resonators with a variable capacitor.

3. I favor size 50 to 80 powdered iron toroids with number 2, 6, or 10 material for a reasonably high Qu.

4. I aim for an insertion loss of of 3-4 dB; especially above the 40 Meter band; *consider the variances discussed earlier*

5. I aim for a return loss of at least 20 dB; *consider the variances discussed earlier*

6. If you can, measure your bread boarded filter bandwidth to confirm or improve the GLPA simulation. Insertion and return loss are easily measured — see EMRFD and the RF Workbench web pages on this site for methods.

I provide no graphic tutorial of DTC08; however, some work flow suggestions follow:

Open up DTC08, choose your center frequency, Qu, inductance and bandwidth and then press the **Calculate** button. Adjust the L until you get close to a standard value coupling capacitor from your parts bin. If required, you may also tweak the bandwidth value to get the needed coupling capacitor. It's wise to change the L before BW since changes in inductance don't cause too many complications within limits.

Name and save your filter to a specific file system directory or folder; or simply save it as the default file.

Open GPLA08 and load your recently saved filter file. Press the **Plot** button and then the **Click to Review Circuit** button. In some cases, you will have to type the CF in the *Cursor Data* text box and press **Plot** to set the cursor at your center frequency.

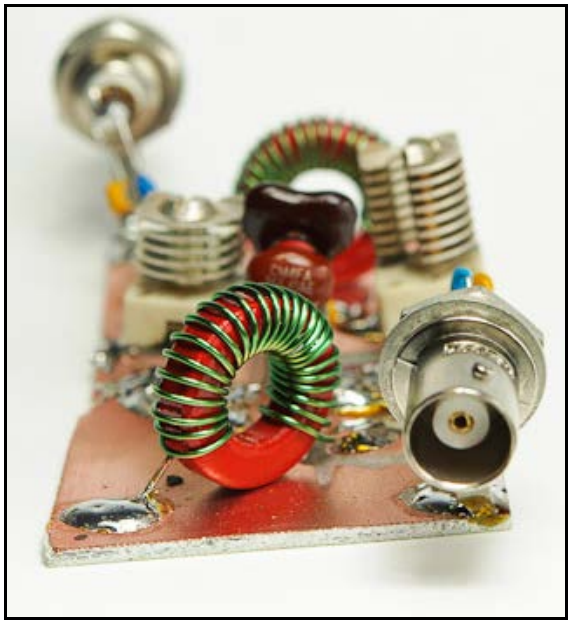
Change the coupling capacitor(s) to a standard value using the **Enter New Value** data entry controls.

Adjust the series end coupling capacitors to standard values. and if S11 is an issue for you, tweak them up and down while observing S11. Re-establish your center frequency by tweaking the parallel tank tuning capacitors and then re-plot to ensure the CF is lined up with the center of the plot.

In Part 3, I provide 3 filter design examples. Your own filter designs will be the most important examples to study.

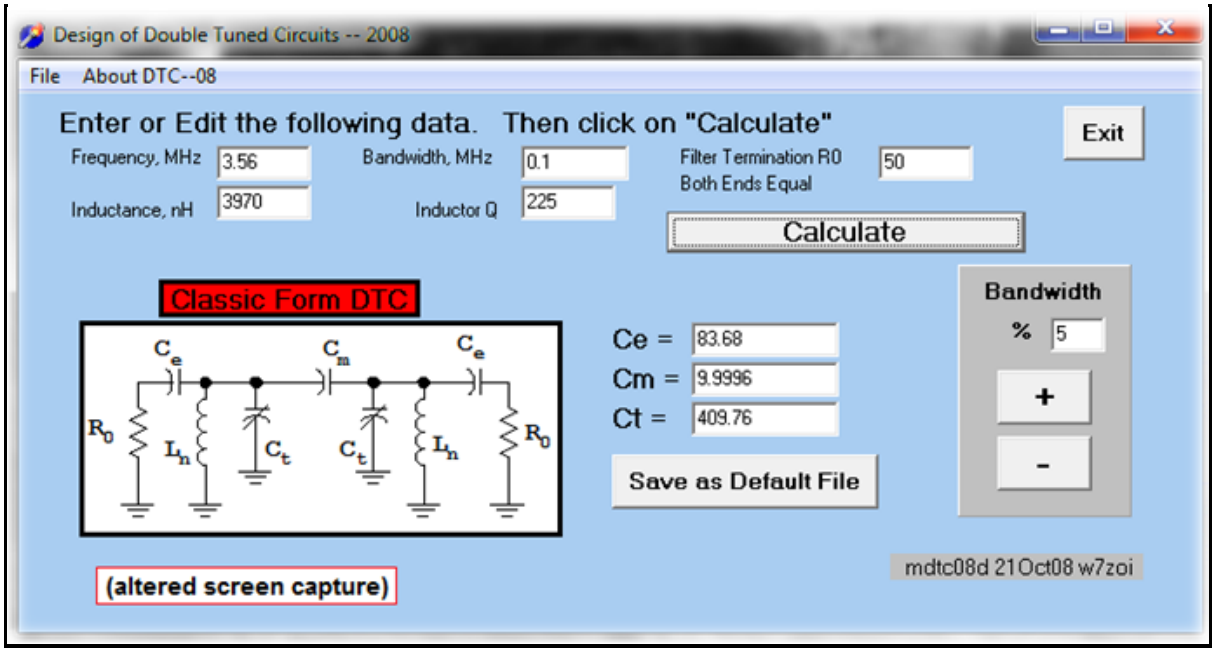
Part 3: Band-pass Filter Examples

Example 1: An 80 Meter Band Filter

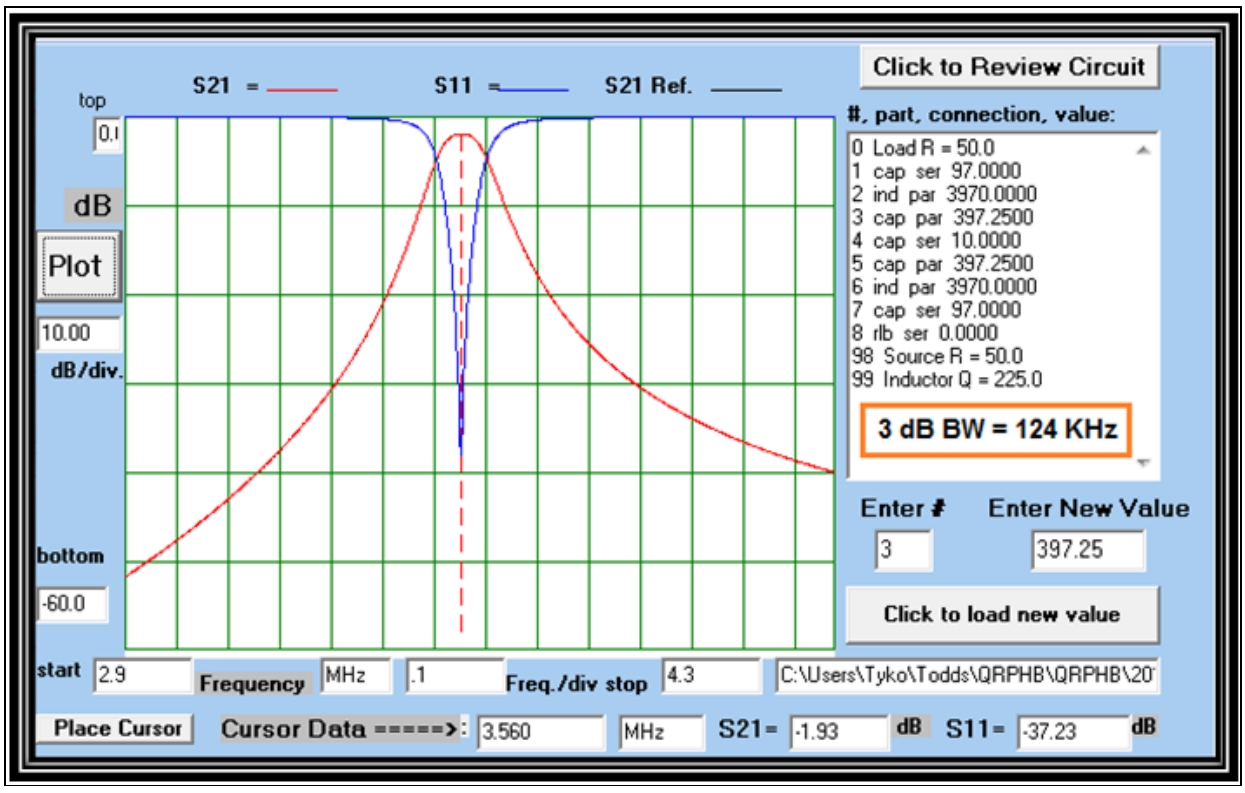


Above — Breadboard photograph of the 80M filter. This example filter may hit home for you — I like listening to CW at and below the 3560 KHz QRP calling frequency, however, another local Ham likes to talk on 75 meters SSB at or above 3790 KHz. This situation calls for a narrow band-pass filter. With my filter, the attenuation at 3790 KHz = ~ 24 dB; had I built a wide bandwidth filter, for example, 350 KHz BW; the attenuation at 3790 KHz, would only be ~4 dB. Perhaps a 3 resonator filter with even steeper skirts would be better?

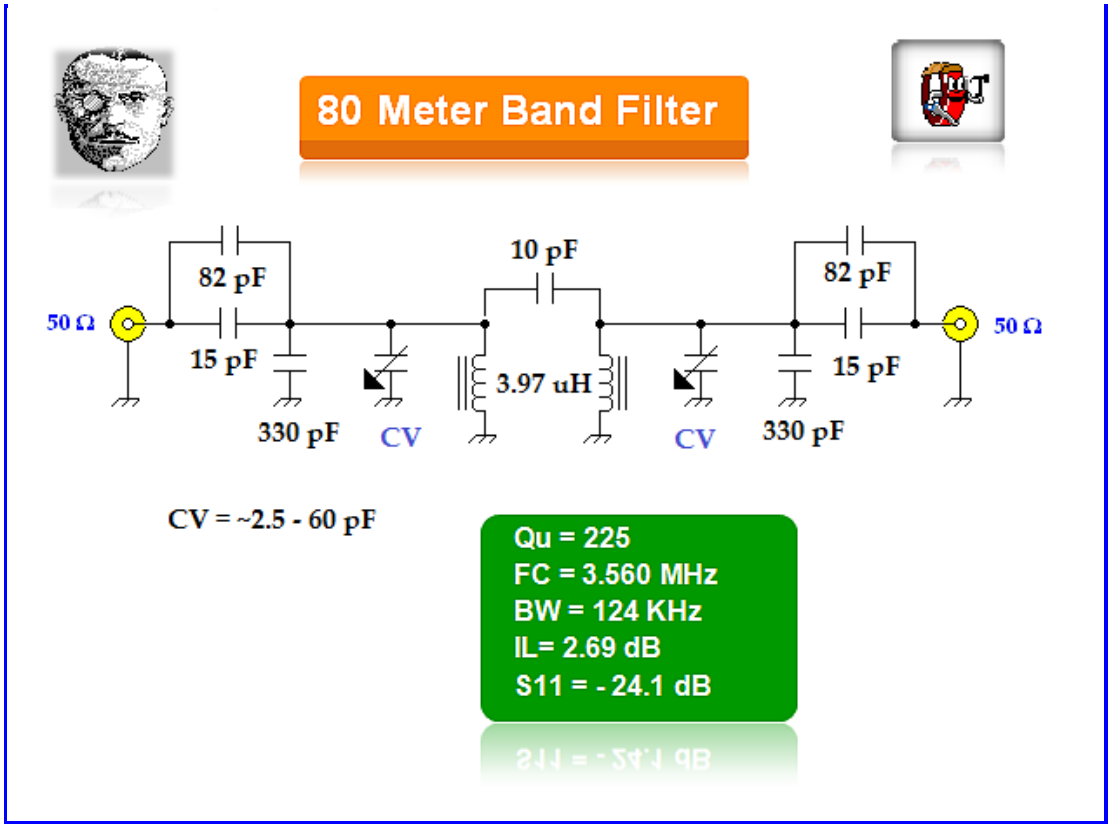
I'll show the design process from start to test.



Above — The basic DTC08 data entry fields were populated. I chose a 100 KHz bandwidth and tried different L values until Cm = 10 pF, since I have a whole drawer of 10 pF capacitors. I believed my Inductor Q would be at least 225 and wound 25 turns of #22 AWG on a T68-2 toroid and expanded or contracted the windings until I measured 3970 nanohenries. In reality, we should measure the inductor Q and in future I will, however, my sense is that few builders do.



Above — After saving my filter, I opened it up in GPLA08 as above. I replaced #4 with a standard 10.0 pF value, and started tweaking the end caps; parts #1 and #7 to gain a better S11 per my obsession with return loss. Retuning #3 and #7 re-establishes the center frequency and allows the S21 and S11 values to be interpreted. I settled on this filter and headed for the bench.



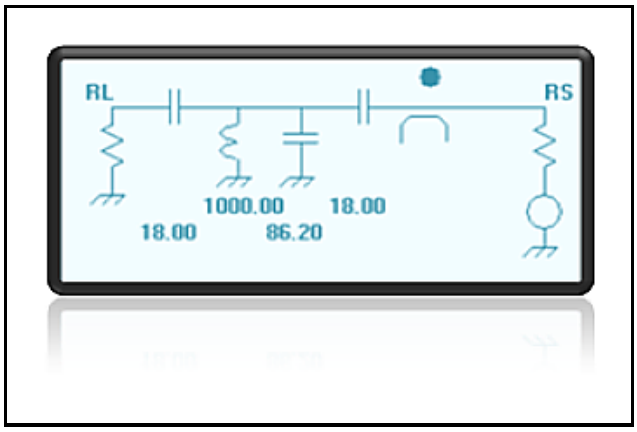
Above — Schematic and analysis of the breadboard. Click for another photograph. In reality, I bench determined the exact capacitance needed to tune each tank at 3.56 MHz with 2 large air variable capacitors that I removed and measured after peaking the filter. For each tank, I try to get just below this value with fixed value capacitors and add a small (2.5 - 22 pF, or so) air variable trimmer capacitor for peaking.

You need to test with capacitance under and over that required to ensure you properly tuned each L-C tank to resonance. Your parts collection, stray capacitance, mistakes or inductance variations in the toroids necessitate custom tuning of your tanks on the bench. I give capacitance values that should work, but it's up to you to ensure resonance of each tank. I find narrow BW filters require a steady hand to tune.

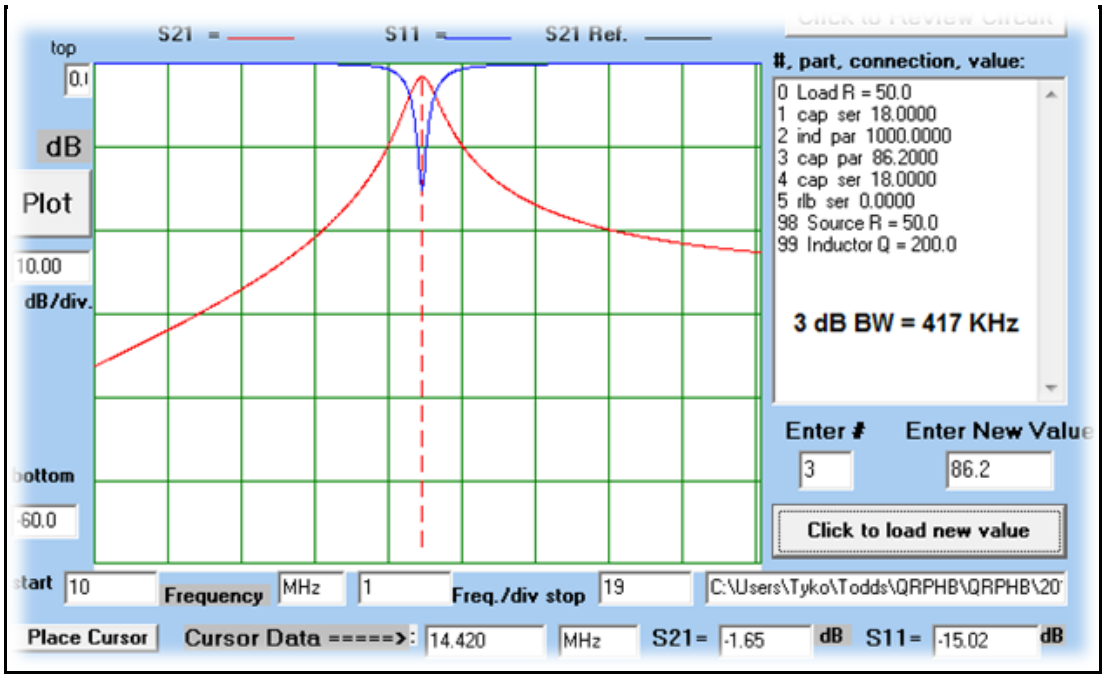
After peaking the tank in your oscilloscope, record the peak-to-peak voltage. Remove the filter and connect your signal generator to your scope with an RF barrel connector and again record the peak to peak voltage. The difference between the 2 is your insertion loss. You can calculate IL with Applet H on this page . Next, perform return loss measurements. If you can, determine the true 3 dB bandwidth of your filter by sweeping it with SA plus a generator. My filter 3 dB BW = 124 KHz.

Example 2: 20 Meter Band Superheterodyne Receiver Filter

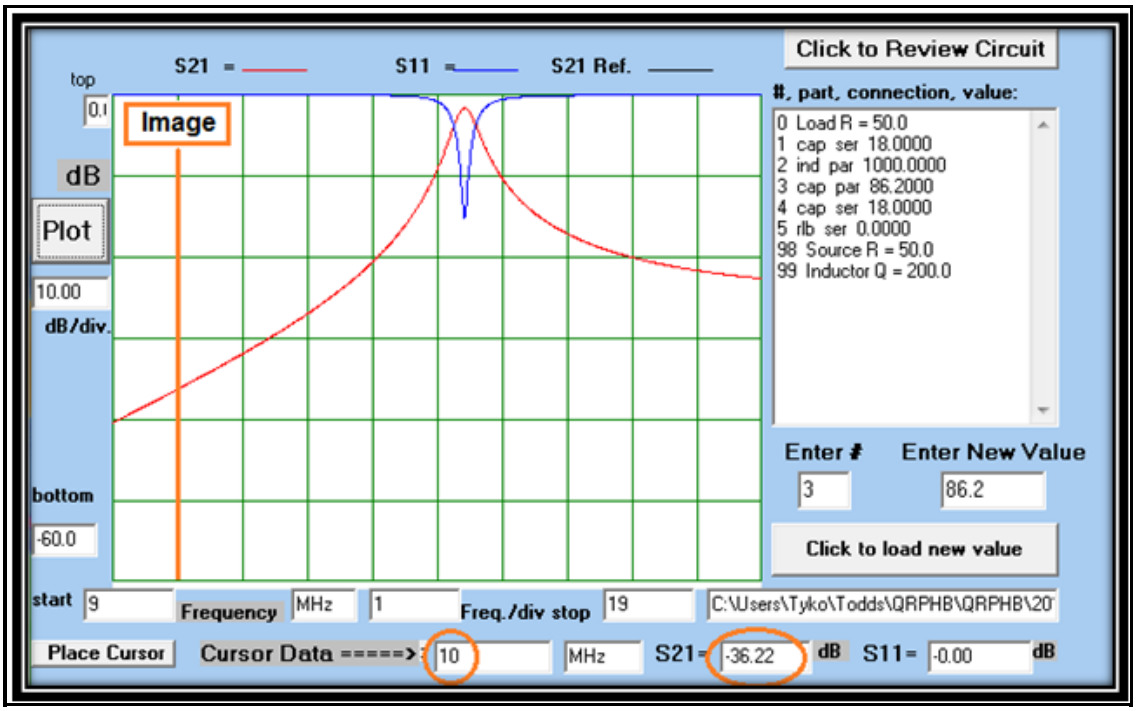
A fictitious builder wants a superheterodyne receiver that covers 14.0 - 14.350. His IF = 2 MHz. The local oscillator = 12 MHz. The image frequency = (12 - 2) = 10 MHz. He centers his filter at 14.020 MHz. In this simulation-only example, we'll go from 1 resonator to 3.



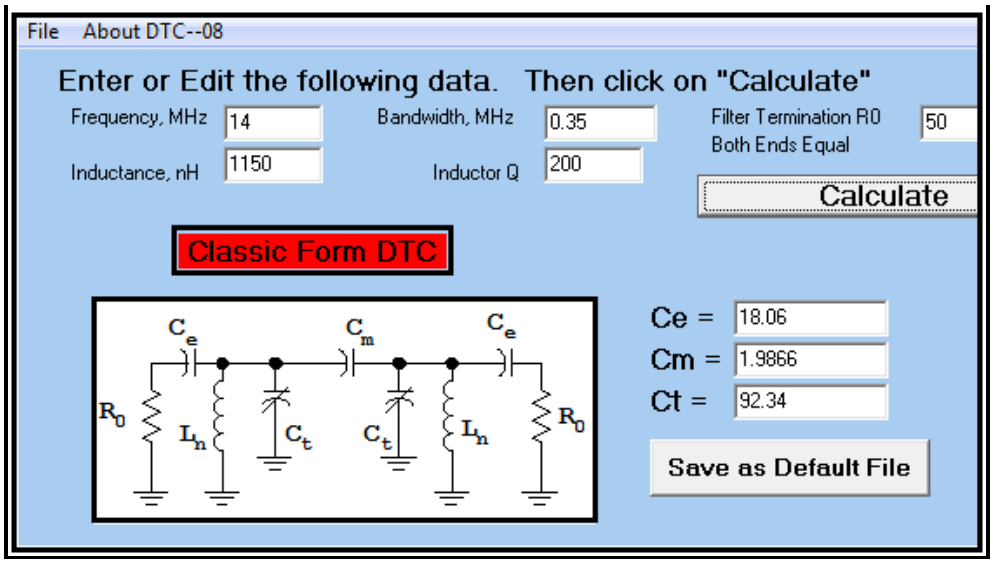
Above — A single resonator with series matching capacitors "built up" in Ladbuild08.



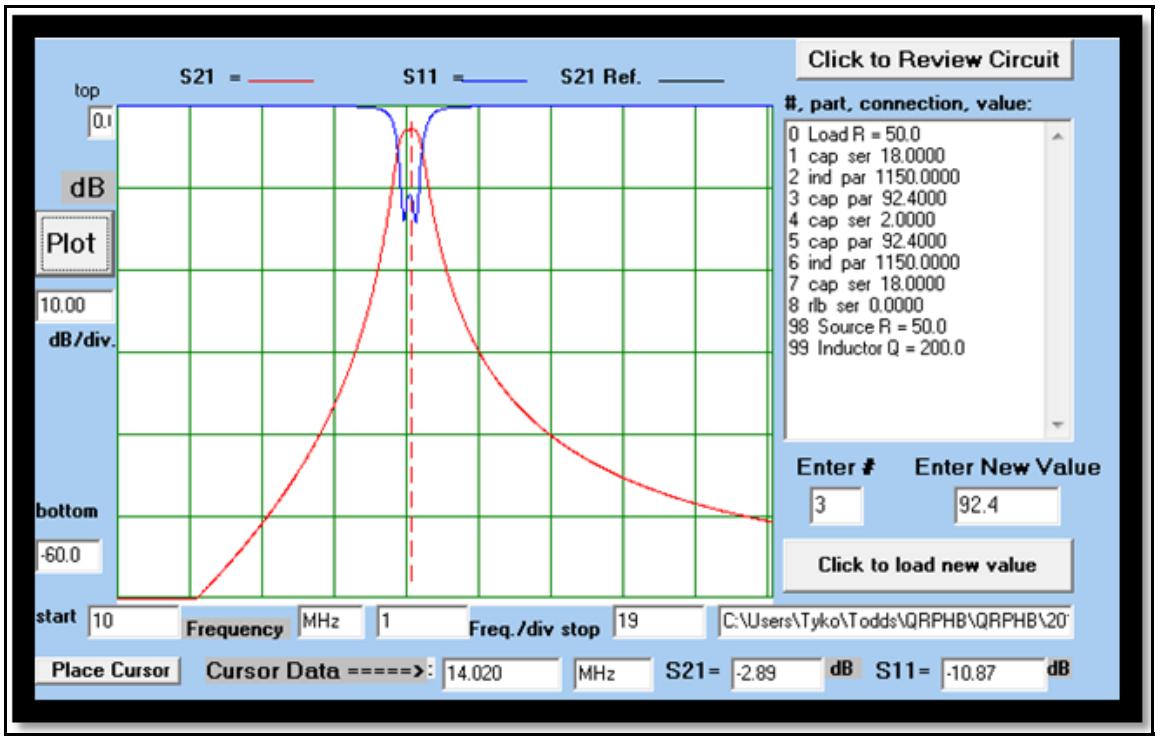
Above — The GPLA08 plot of the single tank filter. The bandwidth = 417 KHz. Increasing the end capacitors to 22 pF to try to increase return loss increases the 3 dB bandwidth as shown here, so we better stick with the original design.
In GPLA simulations with a perfectly centered filter, S21 = the insertion loss and S11 is negative of the return loss.



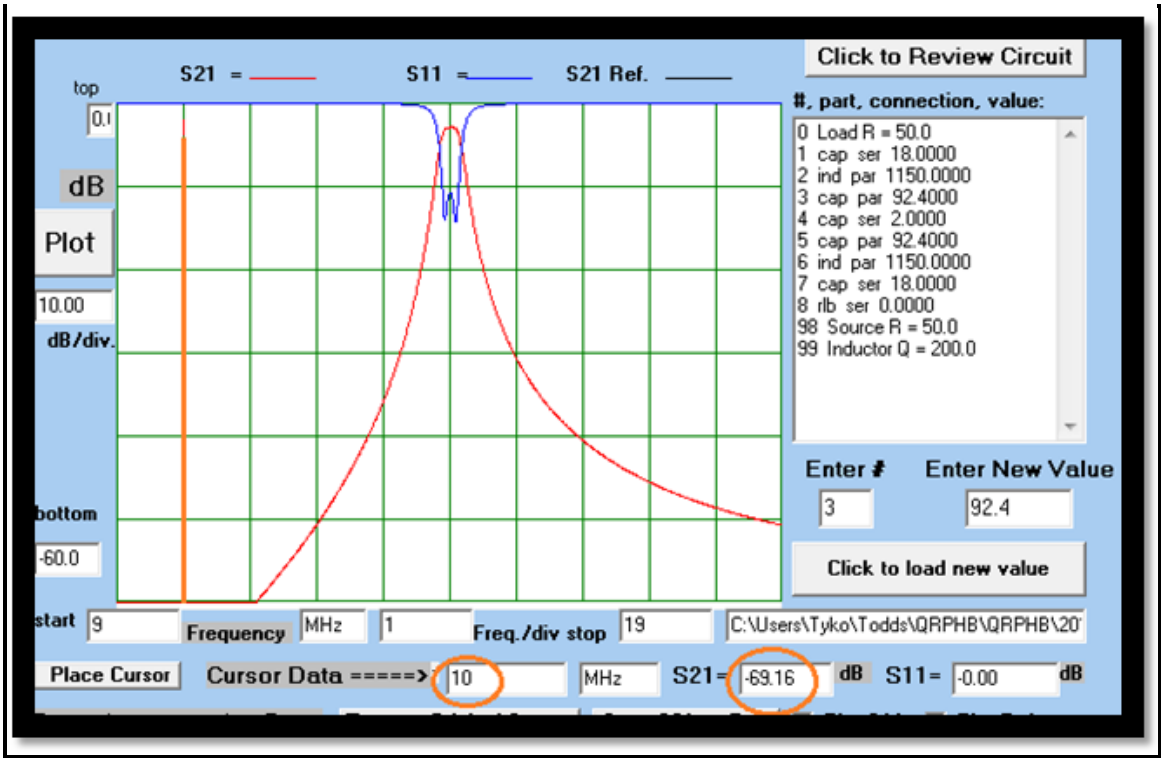
Above — Assessing image frequency attenuation in GPLA08; this sucks — only 36.2 dB down. We need to add a tank.



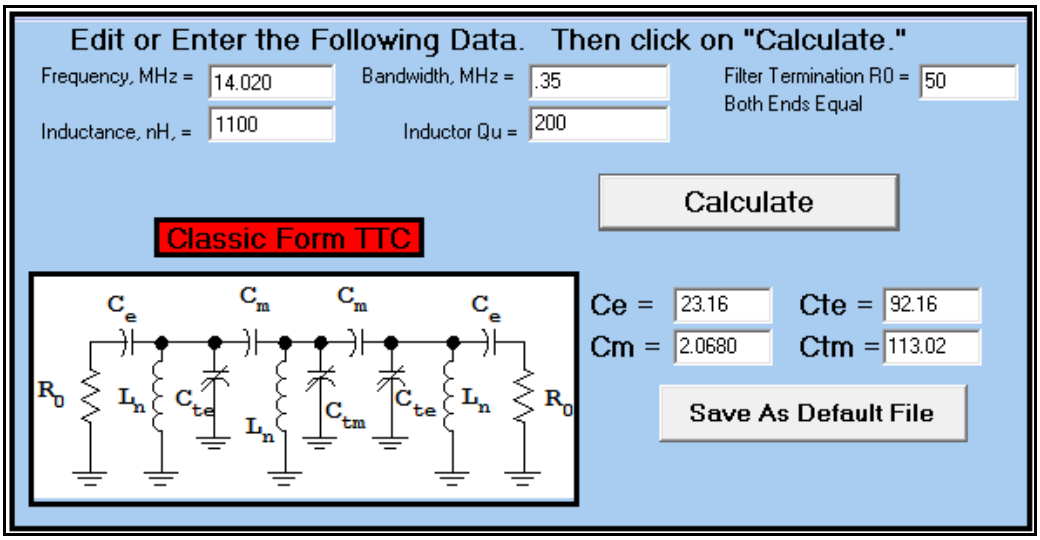
Above — Building up a filter in DTC08, I increased the L from 1000 to 1150 nH to give a Cm near to a standard value.



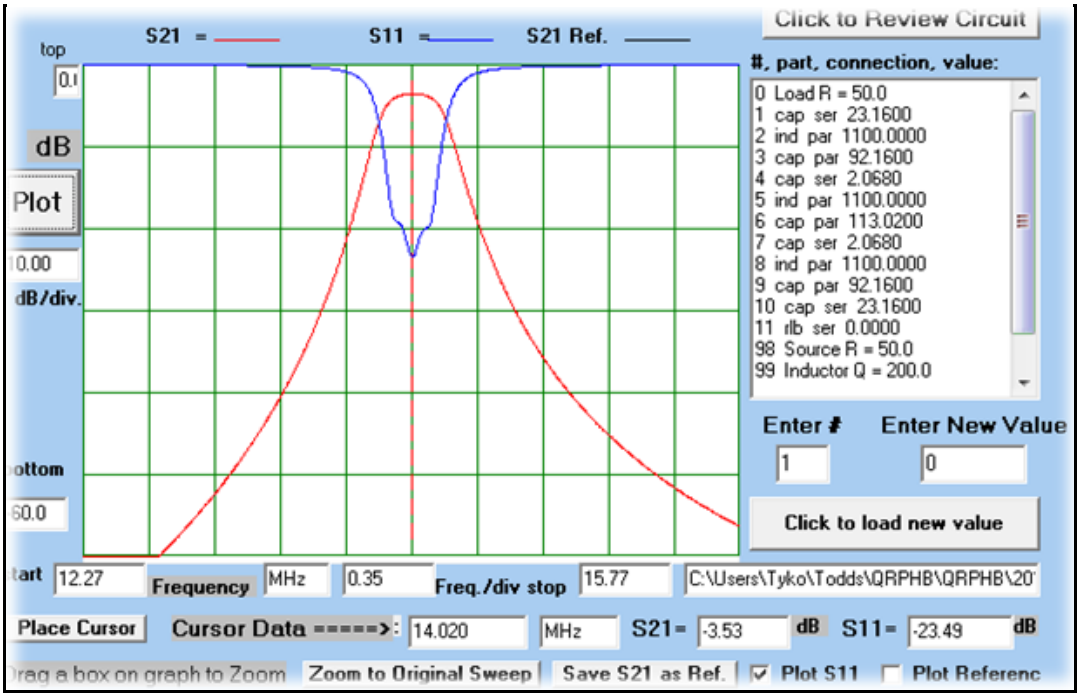
Above — The GPLA08 plot of the double tuned filter. I set #4 to 2 pF and #1 and #7 to 18 pF (nearest standard values). #3 and #5 were slightly tweaked to center the filter. The simulated IL is only up 1.24 dB from the single resonator version. You are probably wondering why I didn't design the filter for a CF = 14.020 MHz in DTC08 above to keep consistency. I probably should have, but wanted to illustrate the versatility of GPLA08 to center filters "on the fly".



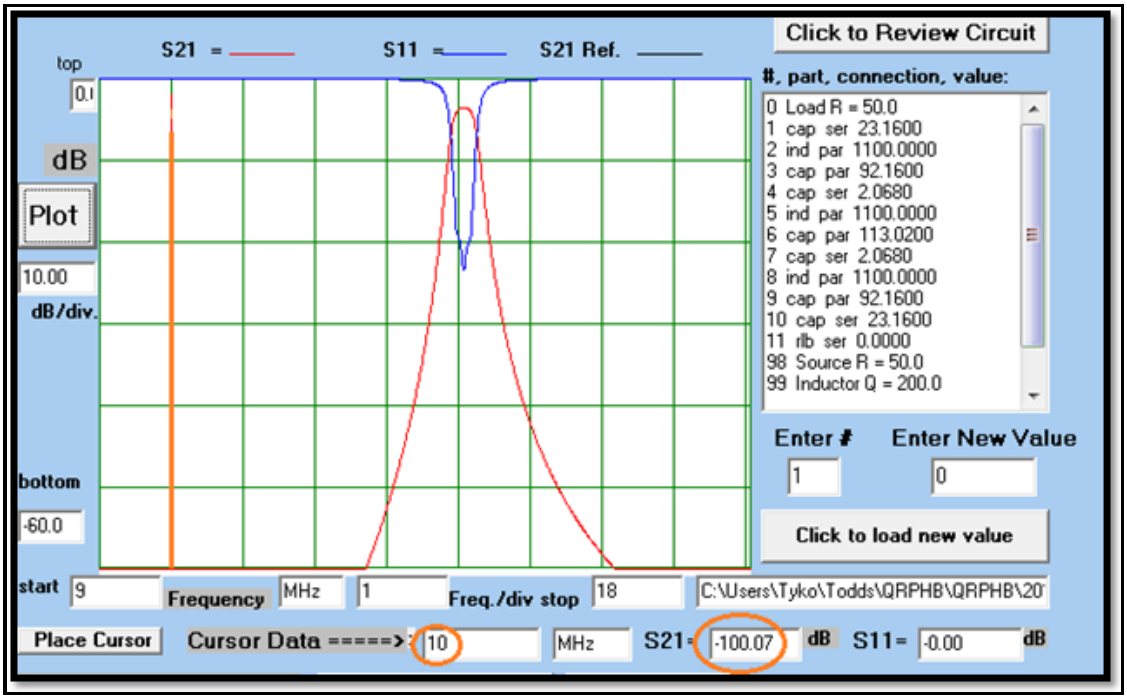
Above — Assessment of the 10 MHz image — now it's 69.2 dB down. Although this filter will work well for his particular receiver specifications, this fastidious builder wants even greater image attenuation and decides to add a third resonator!



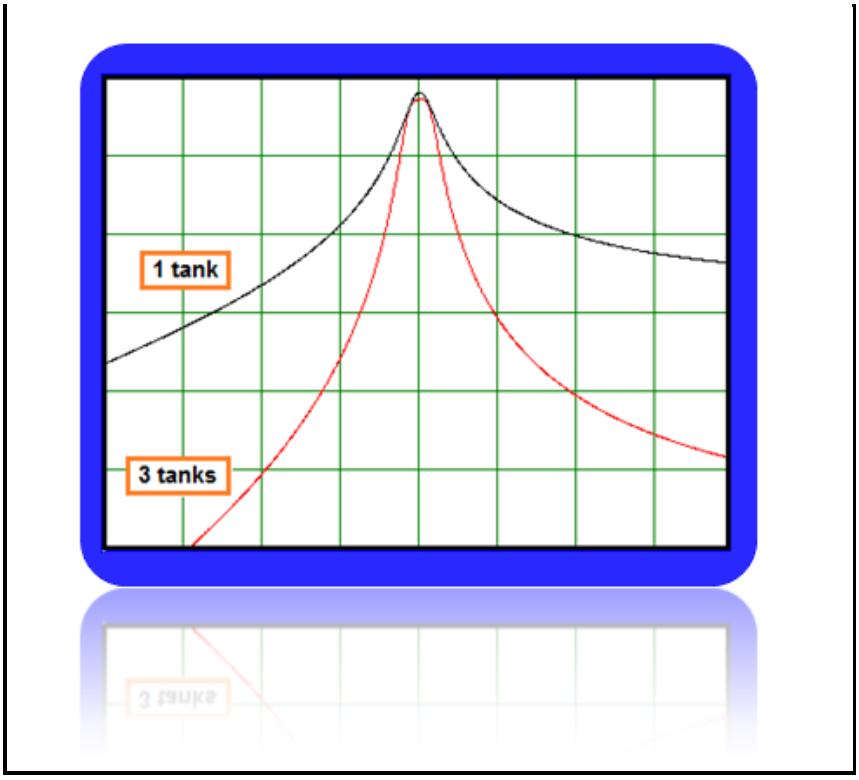
Above — Building up a filter in TTC08, I chose an L of 1100 nH to give a Cm close to a standard value.



Above — The GPLA plot of the 3 tank filter. I performed no parts tweaking — it's up to you from here on in. The simulated IL remains quite reasonable.

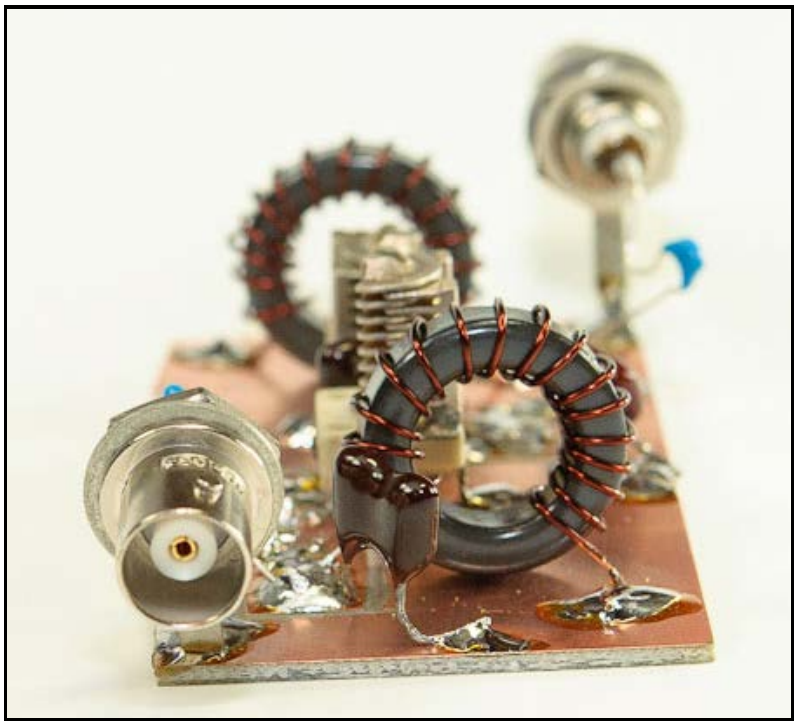


Above — The GPLA08 assessment of the 10 MHz image frequency. Now 100 dB down!



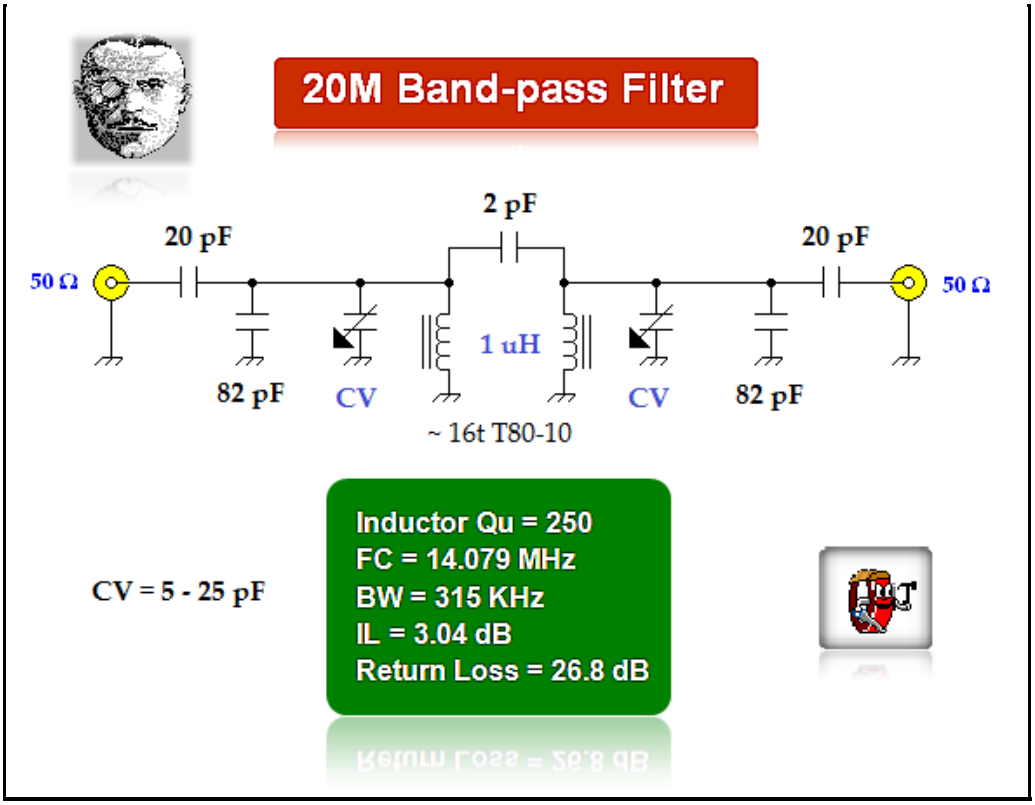
Above — The 1 tank and 3 tank filters superimposed to show the skirt action. The 3 dB bandwidth is the same!

Example 3: A 20 Meter Band-pass Filter for a Builder from Argentina



Above — An Argentinean builder emailed that he wanted a band-pass filter optimized for 14.070-14.095 MHz RTTY but also usable for the CW sub-band and lower SSB frequencies. He wanted a center frequency in the RRTY sub-band and I chose 14.079 MHz. Tuning this filter to a center frequency as low as 7.030 MHz for CW should be possible with the variable capacitor value shown, but as mentioned, you really need to do this carefully on your bench. I employed T80-10 toroids and scrunched or expanded the 16 turns of # 22 AWG wire until they measured exactly 1000 nH.

The best return loss will only occur when your filter is perfectly tuned to the test frequency, so tune carefully.



Above — The schematic + bench analysis for the 20 Meter band double tuned band-pass filter. My original design called for 22 pF series end capacitors to get a decent return loss. After building and measuring the circuit, the results were disappointing: insertion loss = 3.7 dB and a return loss = 17 dB. I wanted a better S11 and IL, so I decreased the end capacitors to 20 pF and savored the measured data shown in the schematic.

Simulating this tweaked design in GPLA08 unveiled a lower return loss than the original design simulation with 22 pF end capacitors; exactly opposite to my bench observations.

Bench work reveals the truth — The filter *you* get is dependent on factors such as parts types + tolerances, stray reactance, layout, test gear and any bench errors. For example, I don't know the Qu of my 1 uH inductors, but suspect that the Qu is greater than the 250 specified. Also my intended - 3 dB bandwidth was 350 KHz, yet my filter = 315 KHz; in part, because I lowered the series end capacitors, but also due to other fore mentioned factors.

Many popcorn builders can't easily measure their filter bandwidth. Does it really matter? Probably not, however, the big realization for me is that unless you measure, you won't actually know your data like insertion loss, return loss, or bandwidth — simulations are great, but don't obliterate the need for bench testing as possible.

Consider this; with the SPICE program you can design a circuit with a 2N3904 and run 400 mA of current through it — the transistor won't smoke 1 bit ! Project outcomes depend on understanding and employing best practices, experience and measurement on the bench. Finding best practices proves difficult in a day and time when general scientific literacy, the number of expert mentors and interest in analog electronics are all waning.

Click for another photo of the filter. On my actual filter, I used high Q, air variable trimmer caps that only had a capacitance variation of 15 pF or so. I soldered in fixed capacitors to get close to the capacitance needed to tune each tank. If possible, I think its better use smaller value trimmer caps because they permit finer tuning. The air variable trimmer offers high Q plus you can see when the capacitor is fully meshed (maximum C). This signals that you need to add more fixed capacitance to that tank for peaking.

Conclusion

To repeat; our parts collection dictates our band-pass filter outcomes. Size 50 to 68 #6 material toroids will work fine for most HF frequencies above 3 MHz. Don't stress out too much if your insertion or return loss is a little higher than you wanted; in all likelihood your filter will work fine and you'll be glad you didn't just copy some else's design and rob yourself of the design experience.

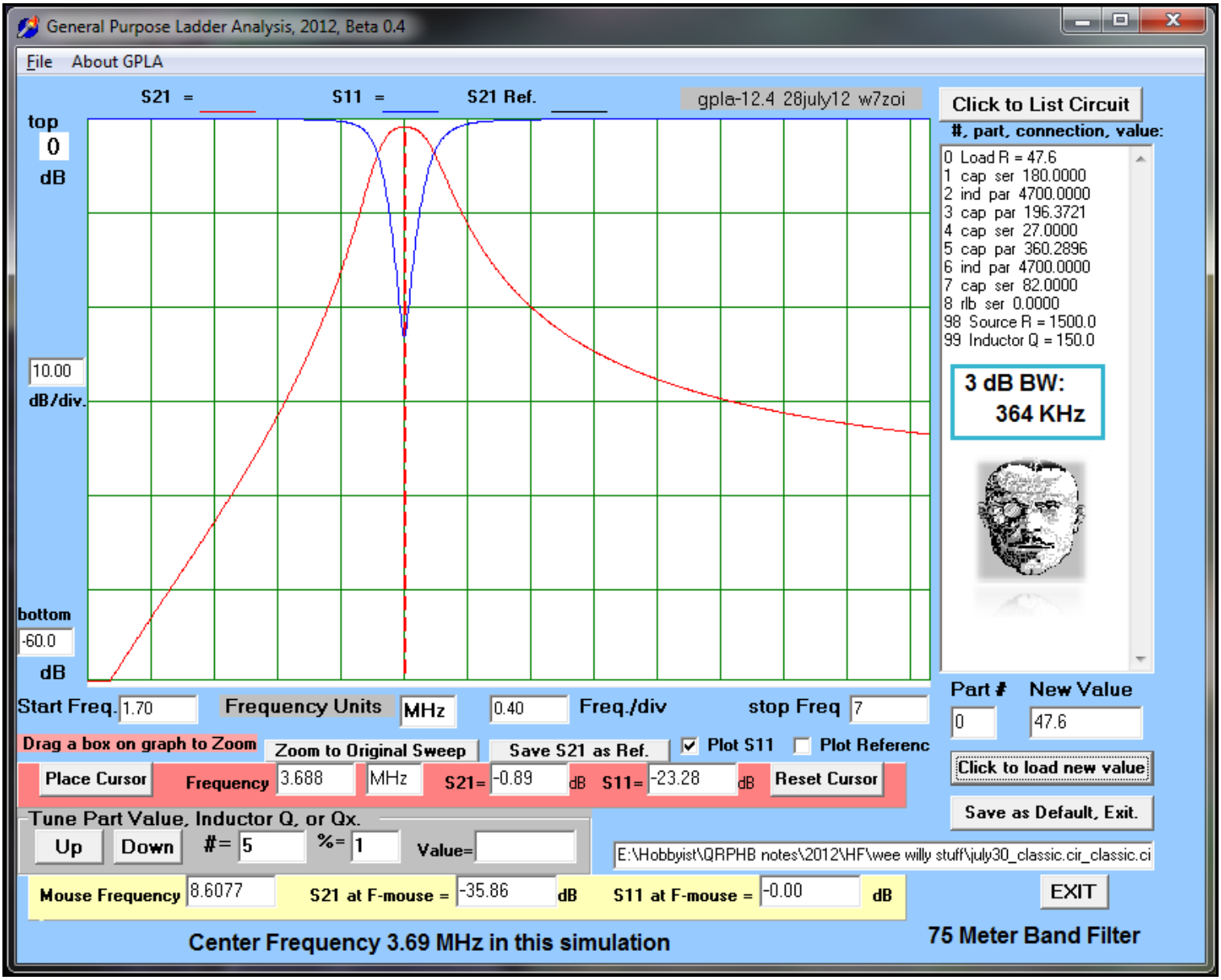
I am hopeful, this web page will inspire a few builders to experiment with band-pass filters for their receivers and other applications. My sincere thanks to Wes, W7ZOI for his guidance with filter design.

QRP — Posdata for August 2012 — NE612 Mixer Band-pass Filters

I designed some band-pass filters for NE612 based front-ends with LadBuild and GPLA and show my 75 Meter band filter design figures below. I'm not a fan of employing an NE612 as a receiver mixer since it easily overloads and spews harmonics when mixing strong input signals. The 1K

'RF gain' pot found in many receivers, or the more conventional switchable attenuator pad prove essential when receiving 'booming' signals with a NE602/NE612 mixer in your front-end.

Still, for field-portable tranceivers/receivers, the NE612 mixer keeps the current and radio size down nicely.

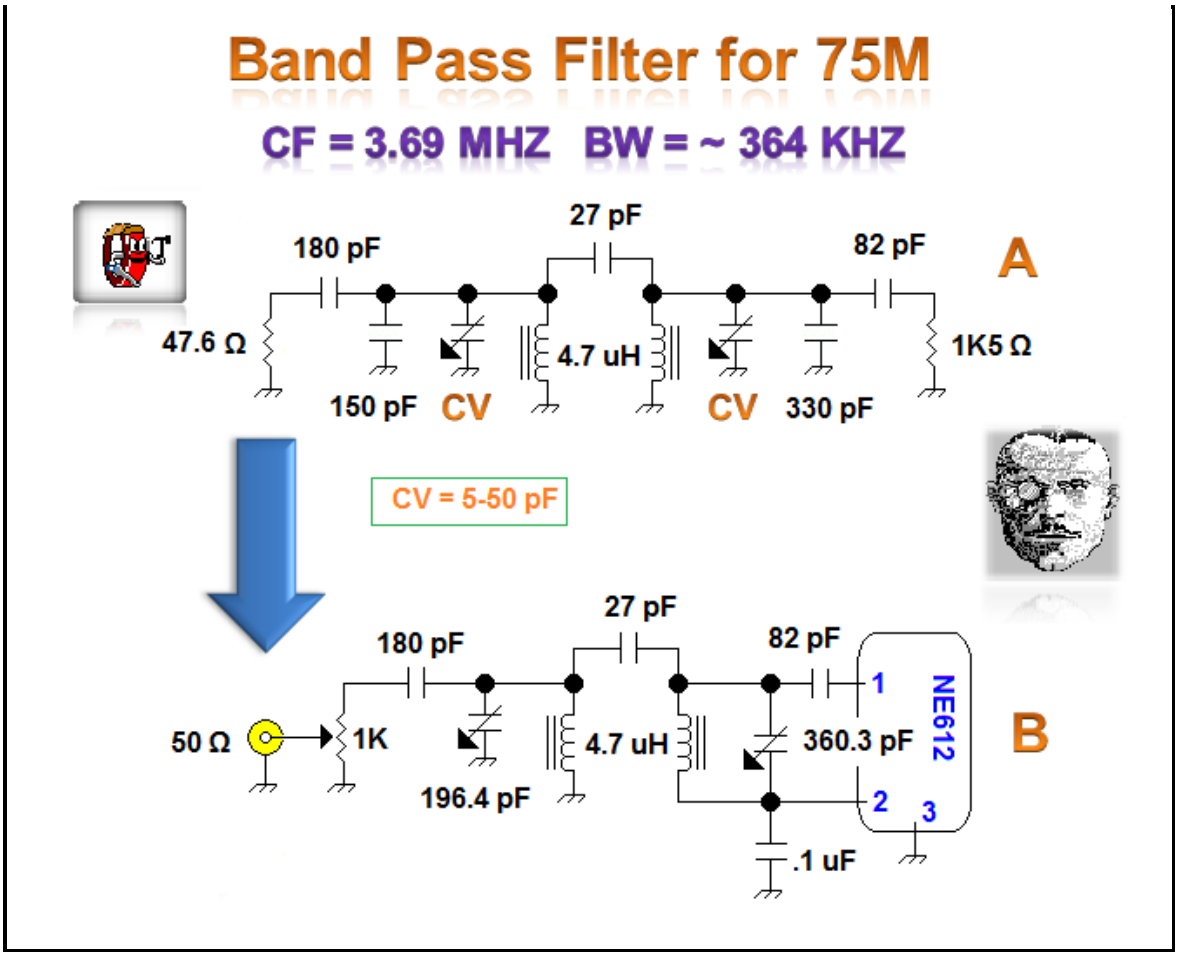


Above — My filter centered at 3.69 MHz. I set the 3 dB bandwidth higher than my usual 200-300 KHz so I could tune a good chunk of the 80-75M band without losing too much signal.

To establish some starting L and C values, I built a classic form filter in DTC08 with a 50 Ω input and output impedance centered at 3.69 MHz. After some tweaking, I settled on L = 4700 nH, Ce = 148 pF, coupling capacitor Cm = 27 pF and about 220 pF (Ct) to resonate each tank.

Next I changed the termination R's to 1500 Ω in DTC08 to simulate the right half of the filter matched into the 1K5 input impedance of an NE612. This gave me some Ct and Ce values to start with. I started Ladbuild 08 and built up a schematic.

In a seperate experiment, I determined that the resonator Qu of a 3.7 MHz L-C tank with an L of 4700 nH wound on a T50-2 core, was ~150.



Above — The completed schematic. My filter exhibits an attenuation of ~70 dB at the top of the AM radio broadcast band (1500 KHz) providing I shield it in an RF-tight box. I took the 47.6 Ω input Z from the 1K pot and my 50 Ω antenna in parallel. An input Z of 50 Ω would work just as well in simulation and on-bench.

In Part A, I show a possible way to resonate each tank with 1 fixed C and a trimmer capacitor. In Part B, I omitted this detail and just show the calculated C needed for resonance as a variable capacitance.

To make this filter with GPLA, I tweaked the capacitor values to nearest standard value parts and tuned the filter with the GPLA Tune Part Value controls while looking at the waveform and my 3 dB bandwidth. I love tweaking values in GPLA and over the years have designed several hundred RF filters for readers.
The rubber hits the road on the bench however! You can get an E.E. degree without melting solder in this day and time — but only bench measurements tell the truth.

Please tune each tank carefully like I mentioned earlier... For example, say a tank needs 180 pF for resonance, *but you don't know this*. You solder in a 100 pF cap and a 5-50 pF *enclosed* ceramic trimmer capacitor into the L-C tank. While watching the 'scope this tank will "peak" since the tank will exhibit its highest peak-peak voltage when the trimmer cap is set to 50 pF and fall off as you decrease the C of the trimmer. You might think you peaked the tank, however you' re actually under by 30 pF!

While leaving the trimmer set to maximum C (peak-peak voltage) in this theoretical example, if you tack solder in another 10 pF cap your 'scope will show an even greater pk-pk voltage. , If you remove this 10 pF cap and then place in a 27 pF cap, the pk-pk voltage will go even higher since you're almost at the target 180 pF. If you removed the 27 pF cap and tack soldered in a 47 pF cap, the pk-pk voltage in the 'scope will go down since your now at 197 pF. Thus you know that resonance is somewhere between 177 and 197 pF.

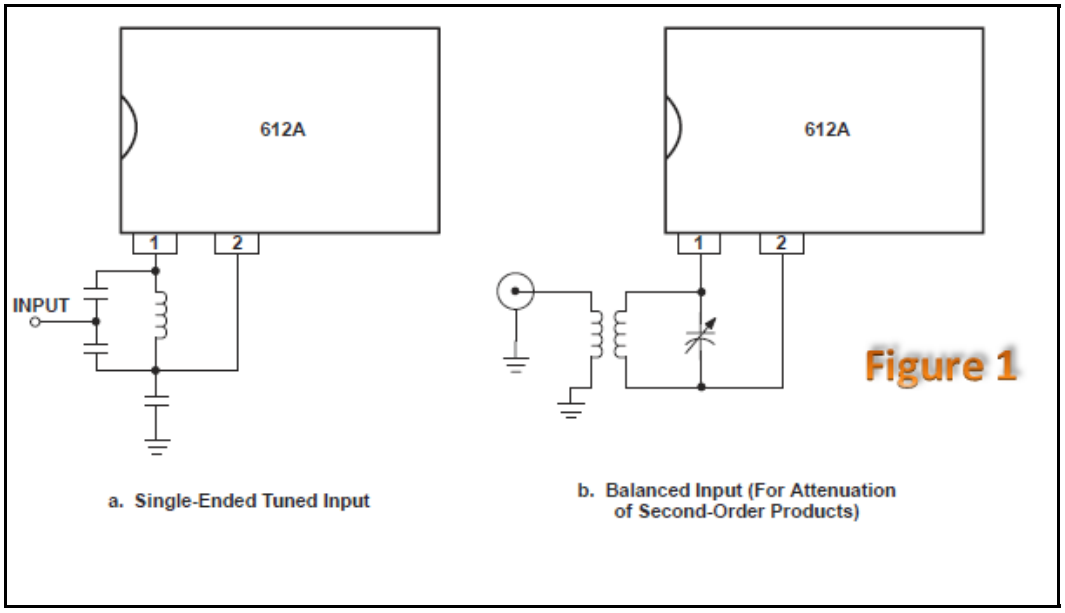
Of course you could decrease the trimmer cap C and stil use the 47 pF cap, however, my description isn't a prescription to follow, just some things to think about. Sometimes I remove a trimmer cap and measure it to ensure the cap is not set to maximum C; that would tell me I need to add more fixed capacitor(s) to the tank. Air variable trimmer caps give visual indication since maximum C occurs with maximum mesh. Unfortunately they are rare and expensive.

On my bench I keep a pair of small 12 to 400 pF air variable caps and temporarily solder them into my tanks. After peaking, I remove and measure them — then I have a good idea of what capacitance is needed to resonate the tank at my test frequency.

It's all an experiment.

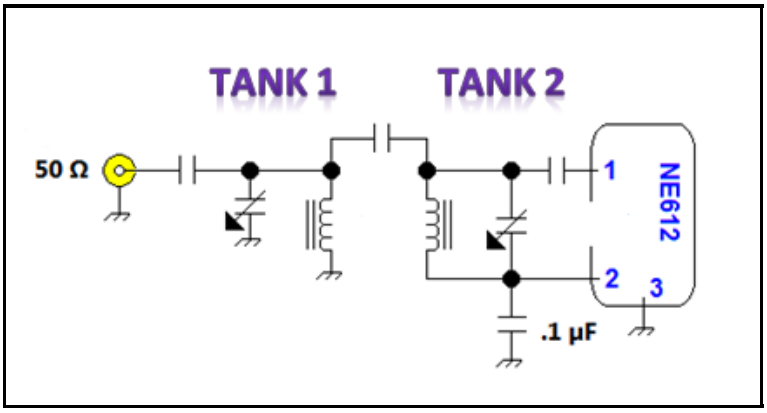
Click for another low loss, well matched example: CF = 5.17 MHz, 3 dB BW = 196 KHz. 50 ohm version. NE612 final version.

QRP — Posdata for August 2012 — More NE612 Receive Mixer Band-pass Filter Experiments



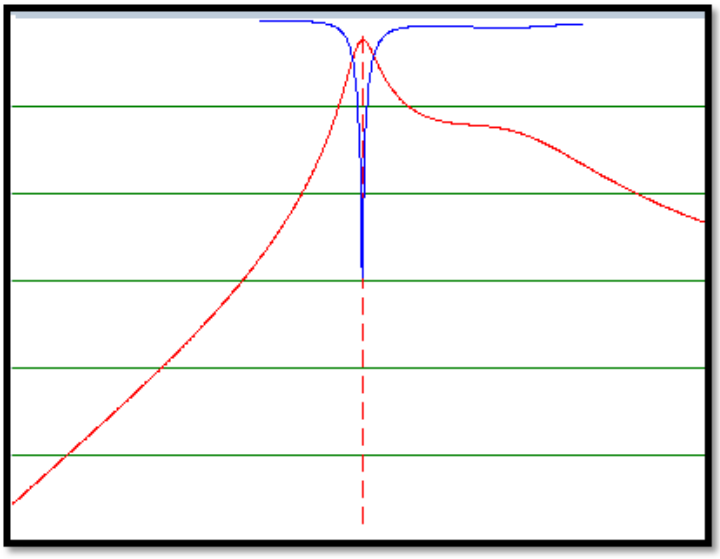
Above — NE612 input circuits. The NE612 datasheet specifies a 3 pF input capacitance + a 1K5 input resistance.

If you look around the Web, many builders just run a single tank for band-pass filtering. While okay for novelty-grade rigs, the poor filter stopband may unleash some ugly problems in the mixer and on down the receiver chain.



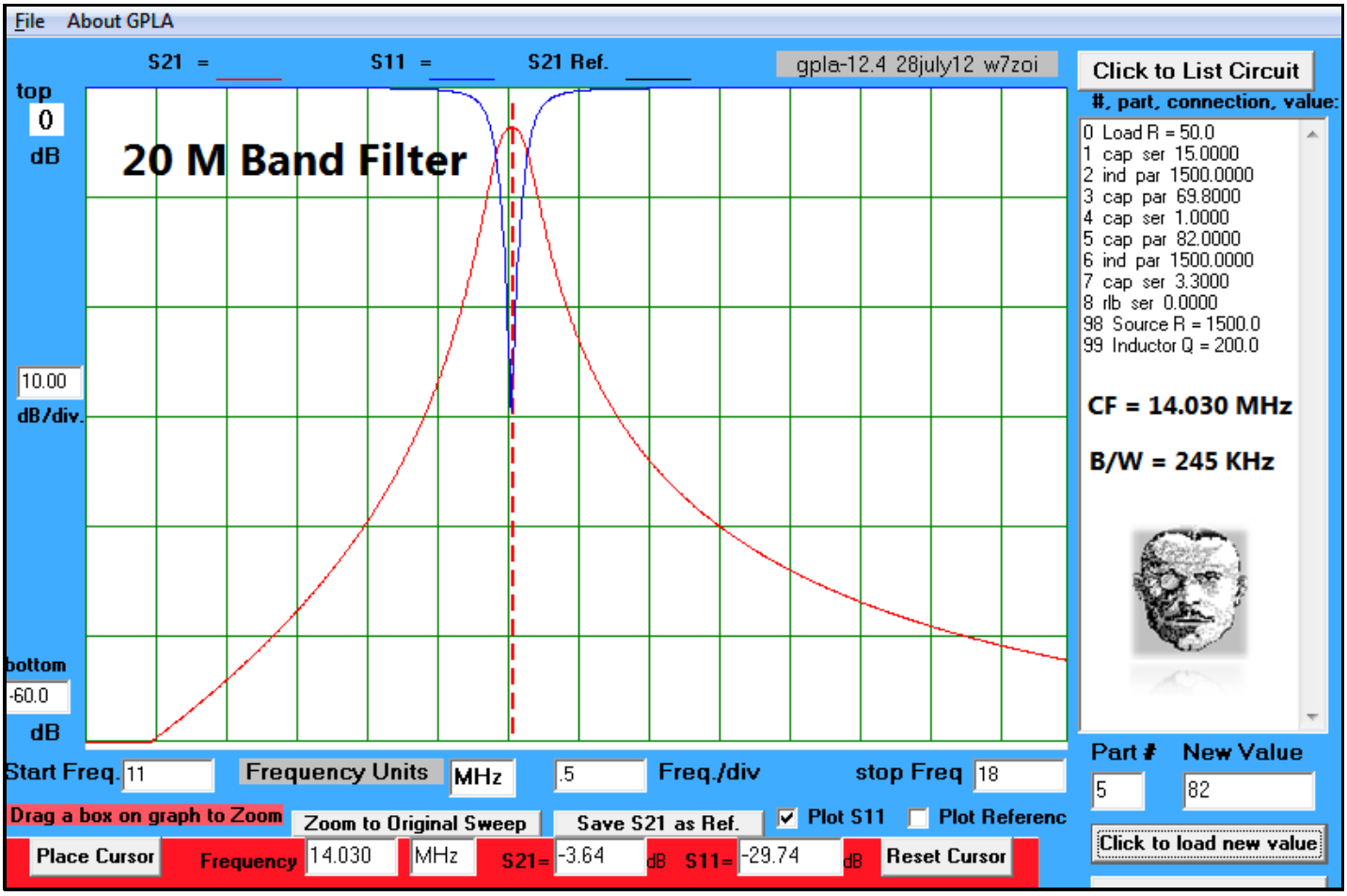
Above — A double tuned circuit with the L-C tanks named 1 and 2 and a series capacitor to match Tank 2 to the NE612.

Most builders match Tank 1 into its 50 Ω source with a capacitor divider, or a matching transformer. For Tank 2, some enthusiasts just connect the Tank 2 coil directly to pin 1 as shown in Figure 1a. Without the matching series capacitor, unfortunate side effects may arise...



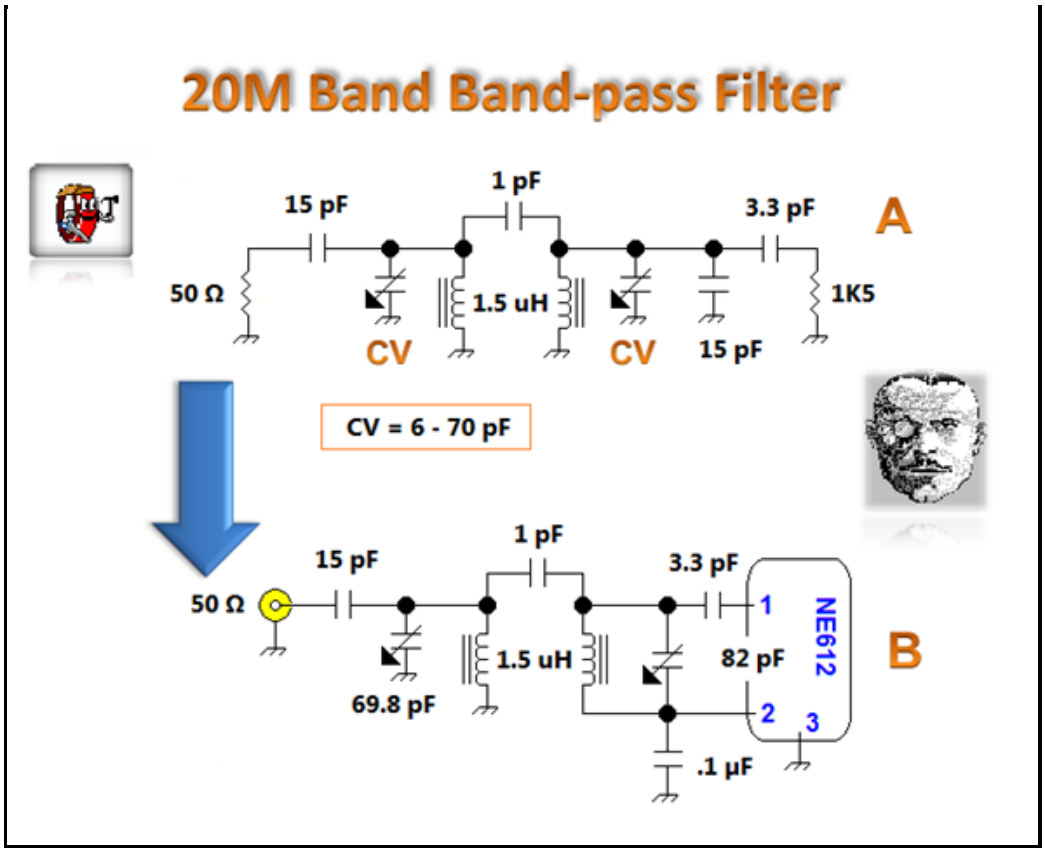
Above — The low-pass skirt of a double-tuned filter may attenuate higher frequencies poorly when no series capacitor (or other network)

matches Tank 2 to the NE612 input. I perfectly matched Tank 1 to its 50 Ω source just connected Tank 2 to a 1500 Ω resistive load in this simulation. I wonder how bad things get in the real world when a complex impedance is involved?



Above — I designed a filter for a 20 Meter band CW receiver centered at 14.030 MHz with DTC and GPLA. The design 3 dB bandwidth = 245 KHz. NE612 filter design was discussed in QRP — Posdata 1.

I then breadboarded the filter with T68-6 inductors, but common lower Qu ceramic trimmer capacitors.

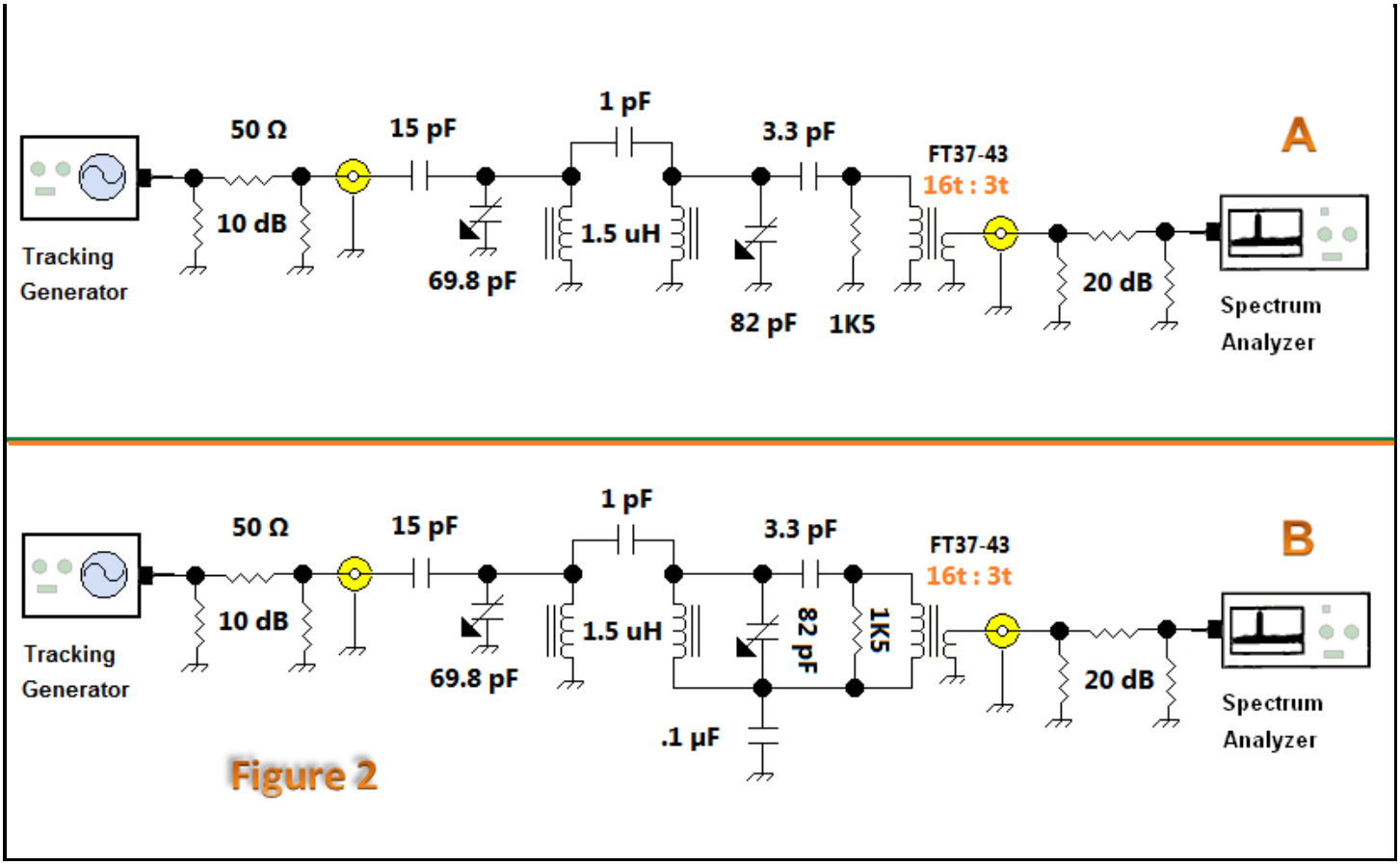


Above — The DTC/GPLA filter design with Tank 2 evolved to provide single-ended input for the NE612.

I wanted to test 2 questions:

1. Does the 0.1 µF coupling cap connected to the cold end of Tank 2 and Pin 2 change the bandwidth or filter skirt shape?
2. Will the 3.3 pF cap really match the NE612?

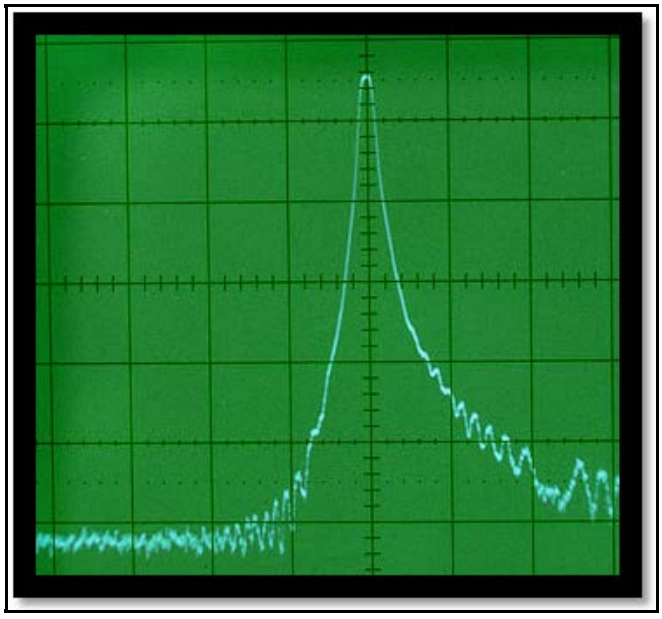
I expect that worldwide, the NE612 input impedance may vary slightly from part to part; different breadboards will exhibit different reactances and that although the datasheet specifies 1500 Ω , we may be dealing with a complex impedance that varies with the aforementioned factors plus perhaps, input frequency. I simply want a good filter with clean skirts.



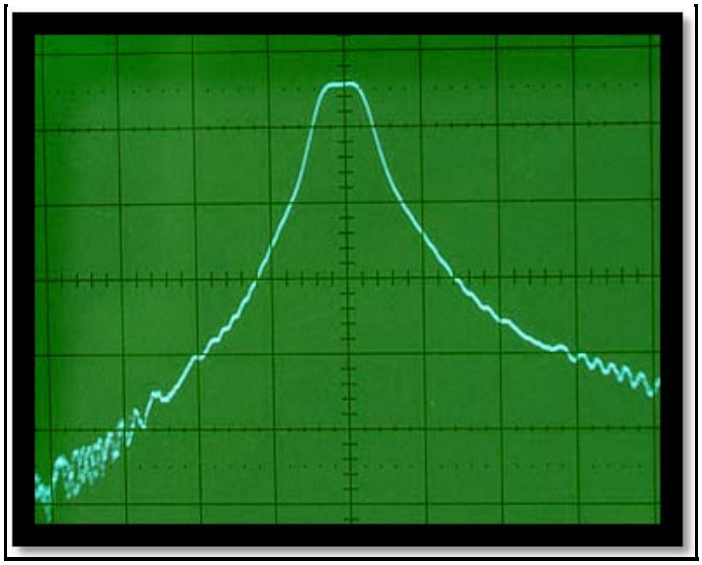
Above — Experiment #1. I compared spectrum analysis of circuit A with B. I built and measured A and then cut away some copper to isolate the copper board grounding the Tank B parts. This "island" was AC coupled to the rest of the ground plane with a short leaded 0.1 µF ceramic capacitor.

I saw no significant difference between Circuit A and B — it appears the 0.1 µF capacitor does not affect the filter parameters to any extent.

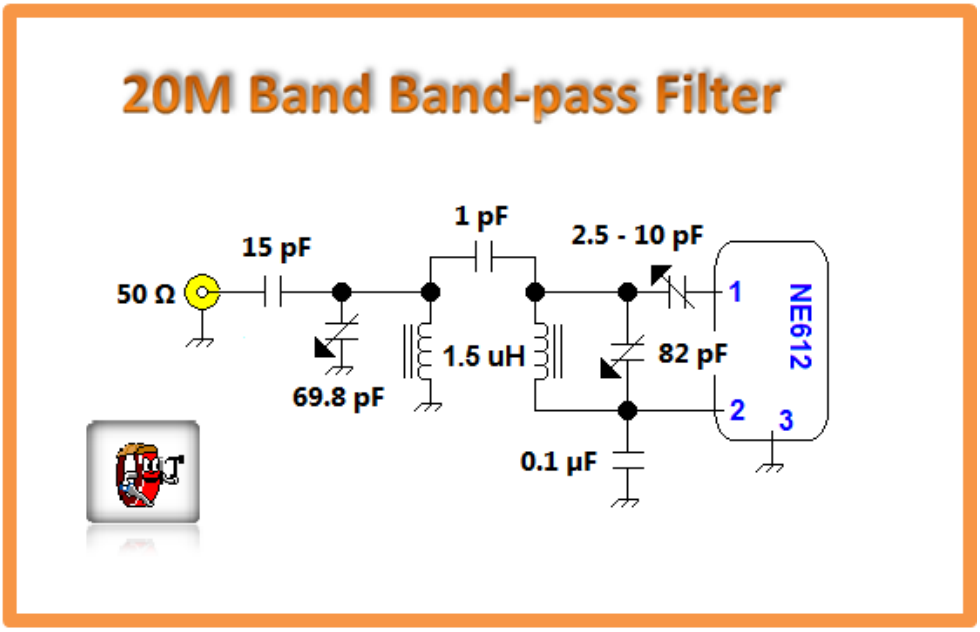
To simulate a NE612, I soldered a 1K5 Ω 5% resistor across Tank 2. Tank 2 was transformer coupled to the 50 Ω Z required for spectrum analysis. A 20 dB pad ensured a strong return loss and a safe input amplitude for the SA.



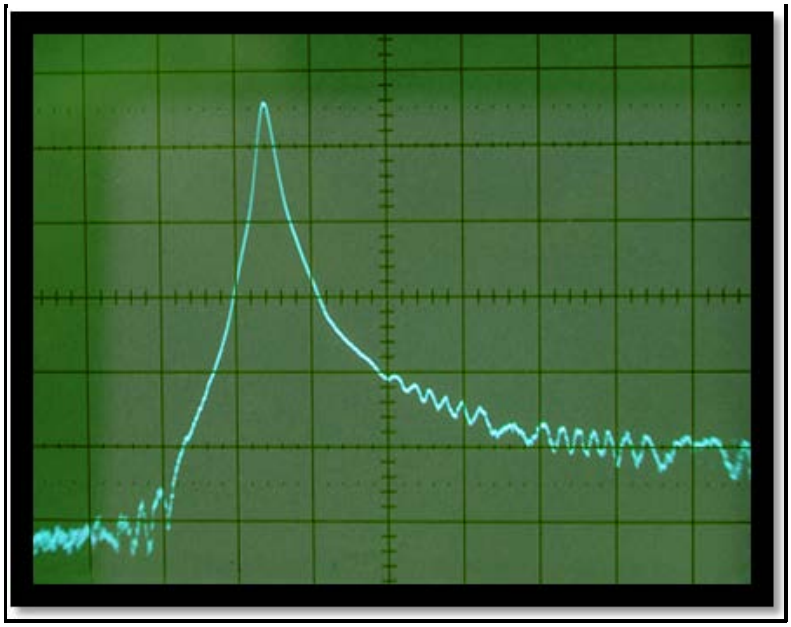
Above — Spectrum analysis of Figure 2A or B. I saw flat topping of the waveform – almost double humping with a higher than wanted 3 dB bandwidth; this bothered me. Fixing this problem was Experiment #2.



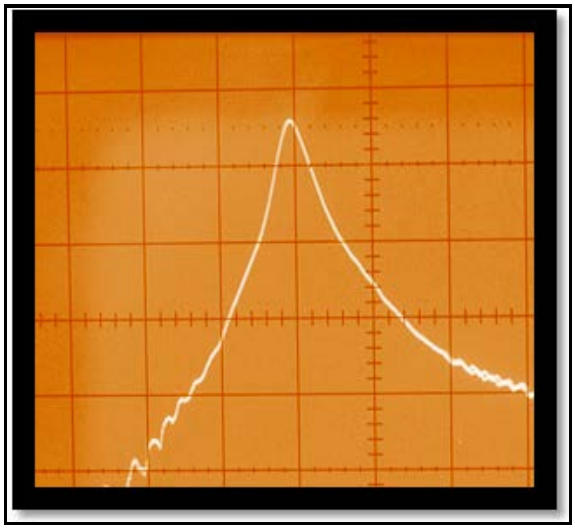
Above — A zoom of the poor coupling of Figure 2A or B. From my experience building filters, I suspected a termination resistance mismatch in Tank 2 ; exactly what I was trying to avoid!



Above — I swapped a trimmer capacitor for the series 3.3 pF fixed cap in Tank 2. Then I reconnected the circuit into my test set up.



Above — A photo of my TG + SA measurement after tweaking the newly added trimmer capacitor and peaking each tank. I shifted the SA screen center over so the tracing could be seen without all the hash marks in the center. Wonderful.



Above — A zoom of the now matched Tank 2. Due to the bench-altered Tank 2 match, my low-Q variable capacitors and other factors, the 3 dB bandwidth now is just over 300 KHz.

I'm not sure if these experiments reflect what actually happens with a NE612 input band-pass filter, however, I plan to match my second tank with a series trimmer capacitor in future NE602/NE612 work. I also want to explore balanced input.

RF — Test and Measurement

RF Workbench Page 4

The 4th installment of a QRP/SWL HomeBuilder series exploring *basic* RF measurement

Part 4 describes a method to calculate reverse isolation in the **50 Ω environment** after converting measured peak-to-peak AC voltages to dBm. I tested 2 common amps at ~7 MHz to show the concepts and calculations.

In this series, I gratefully borrow from the work of Wes, W7ZOI per correspondence, direct contributions and from EMRFD.

Tools Needed

1. 50 Ω terminated scope (or a spectrum analyzer) and a 50 Ω signal generator
2. 50 Ω RF cables with RF connectors (such as short cables with female BNC connectors)
3. 6 dB 50 Ω attenuator pad; plus an adjustable attenuator if you use a fixed output signal generator.
4. BNC through-response connector(s)

Procedure

1. Measure the amplifier forward power gain. This is S21.
2. Measure amplifier reverse power gain. This is S12. Like S11, express S12 as a negative value.
3. Reverse isolation using dB values = S21 - S12.

Step 1 : Measure the forward power gain

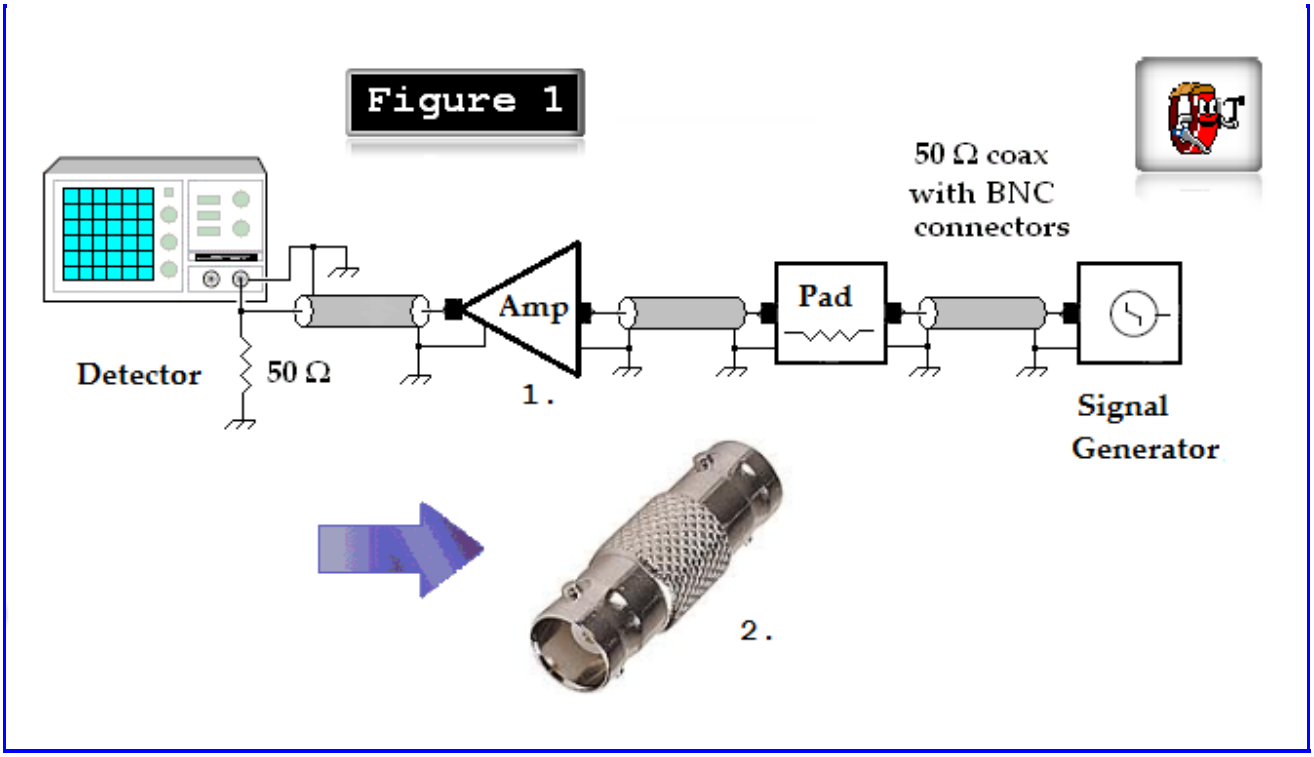


Figure 1 shows how to measure forward gain in a 50 Ω environment — **1**. Convert the measured AC voltage to dBm, **2**. Disconnect the amplifier, insert a through-response connector and convert this measured AC voltage to dBm. The difference between the 2 values = S21. Applet H will do these calculations from the peak-peak voltages. *The attenuation pad following the signal generator in Figures 1 and 2 signify that the signal generators have a 50 ohm output impedance and is optional.*

Choose a signal generator level that ensures the output of your amplifier is linear while providing a good signal to noise ratio for measurement. With an oscilloscope, I generally test amplifiers with an input power of between 0 and -11 dBm; although choose whatever level that works for you consistent with linear amplification.

After measuring the forward gain of your amp, a good way to test for linearity is to add a fixed 6 dB pad between your signal generator and your amplifier to drop the applied signal by half. *[A 6dB pad drops the peak-peak voltage by 1/2 . A 3 dB pad drops the power in dBm by 1/2]*. The power gain should be equal or nearly equal to the measured dB value obtained before you added the 6 dB pad. If they vary significantly, you are likely driving the amp too hard and causing some non-linear output products. S21 = complex *linear* gain.

Step 2 : Measure the reverse power gain

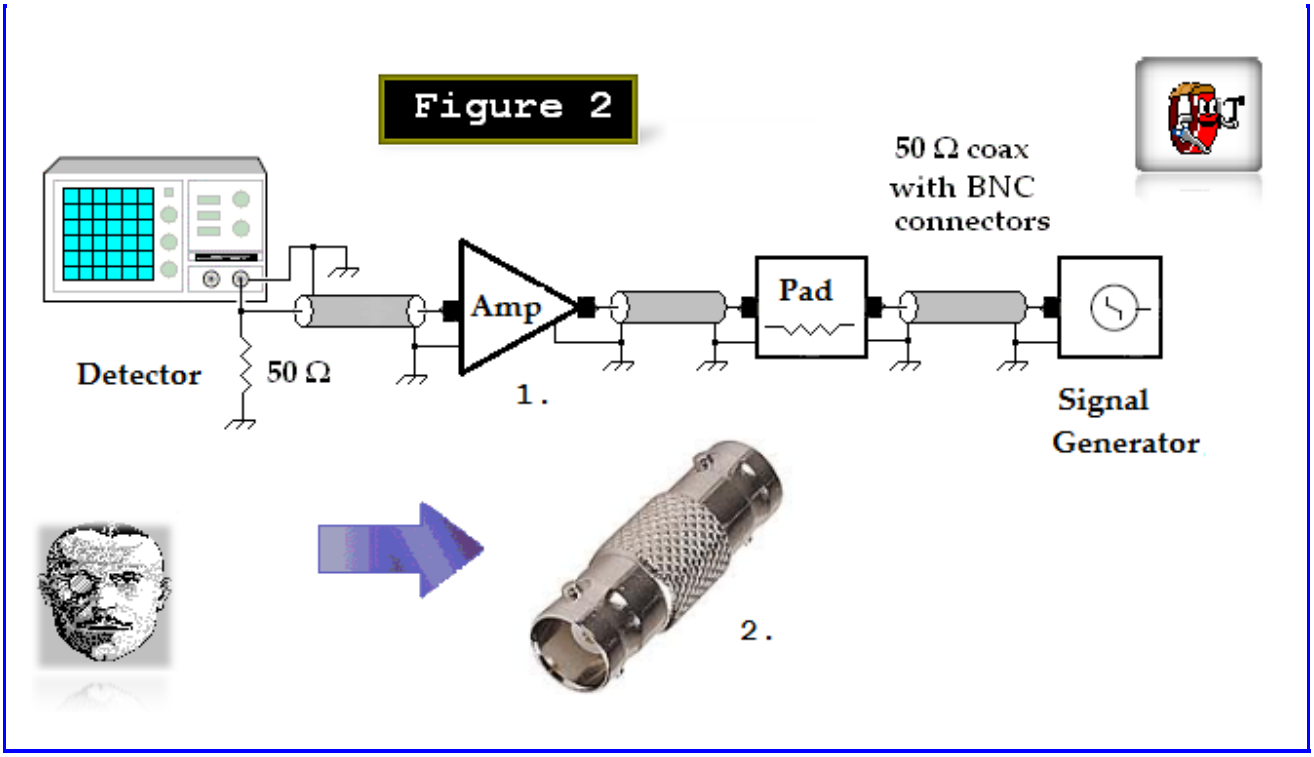


Figure 2 shows how to measure reverse gain in a 50 Ω environment — **1**. Convert the measured AC voltage to dBm, **2**. Disconnect the amplifier, insert a through-response connector and convert this measured AC voltage to dBm. The difference = S12. Ensure that you employ the same drive level used to measure the forward gain.

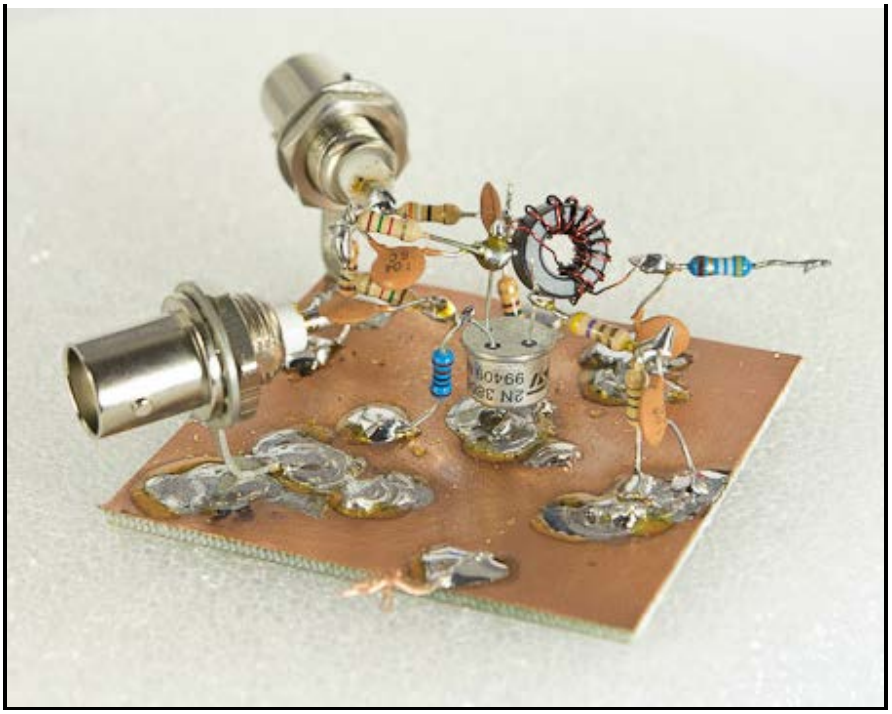
Measuring reverse gain may be tough. When the amplifier under test requires a low drive level and/or has strong reverse isolation, you may not have enough signal to accurately measure with your oscilloscope. The tool of choice for low reverse voltage measurment is a spectrum analyzer (SA) — a narrow band SA may be required to distinguish the weak signal from random noise with low signal voltages.

Summary

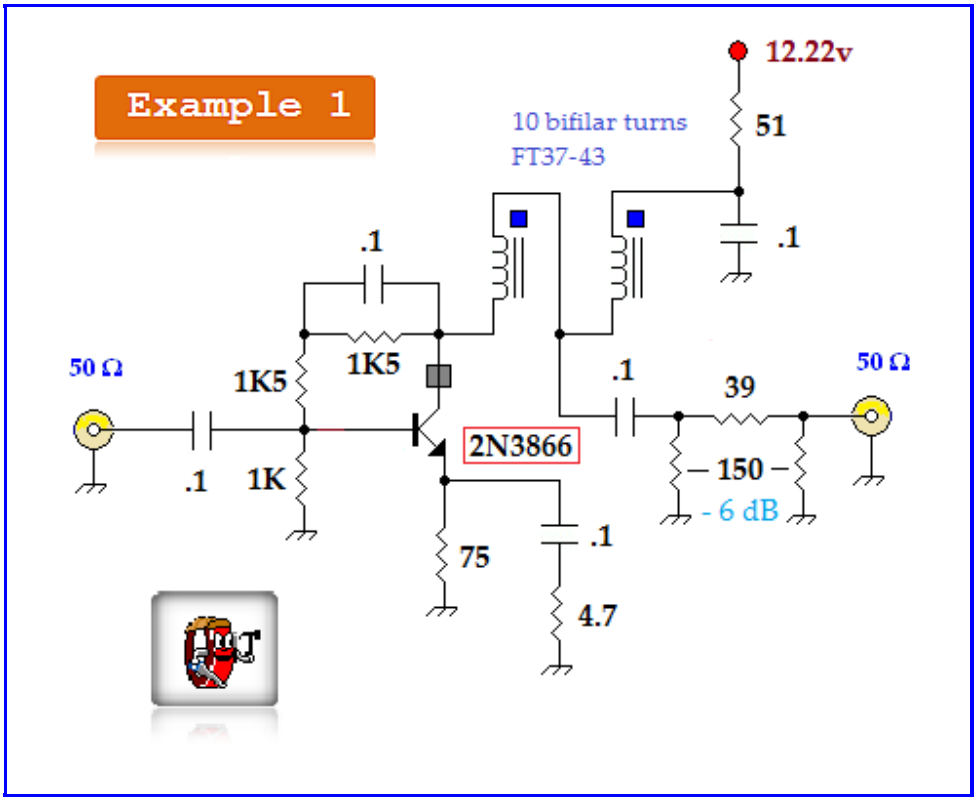
A **practical** bench work flow goes something like:

1. Measure the through-response peak-peak voltage.
2. Measure and record the peak-peak AC voltage while driving the amplifier input port (forward gain set-up).
3. Reverse the amp so you're driving the output port and then measure the peak-peak AC voltage (reverse gain set-up).
4. Calculate S21 and S12.
5. Calculate reverse isolation.

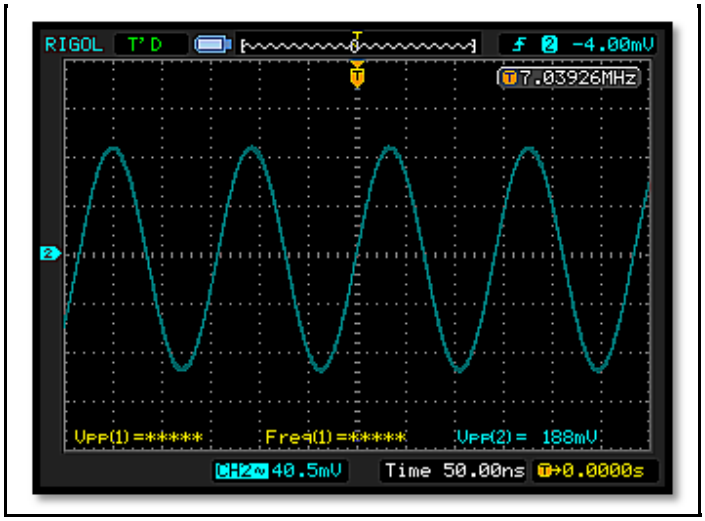
Example 1: Feedback amplifier



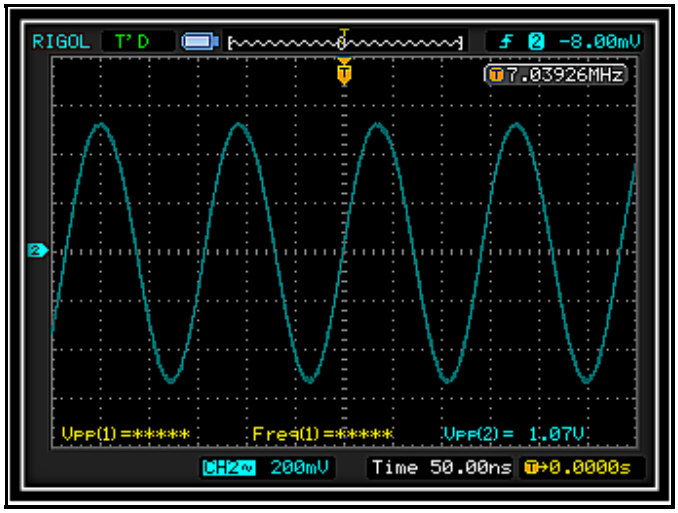
In the first example, I measure the reverse isolation of a Beaverton Special feedback amp. The schematic is shown below. Tested with a 7.039 MHz signal generator possessing a return loss of 30 dB.



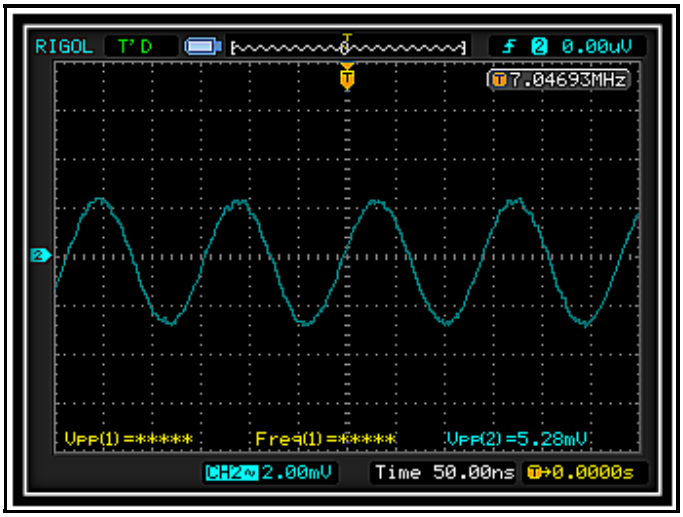
50 ohm Voltage Measurements:



Above — AC voltage with a through-response connector in-situ: 188 mV pk-pk



Above — AC voltage with forward gain set-up: 1.07 V pk-pk



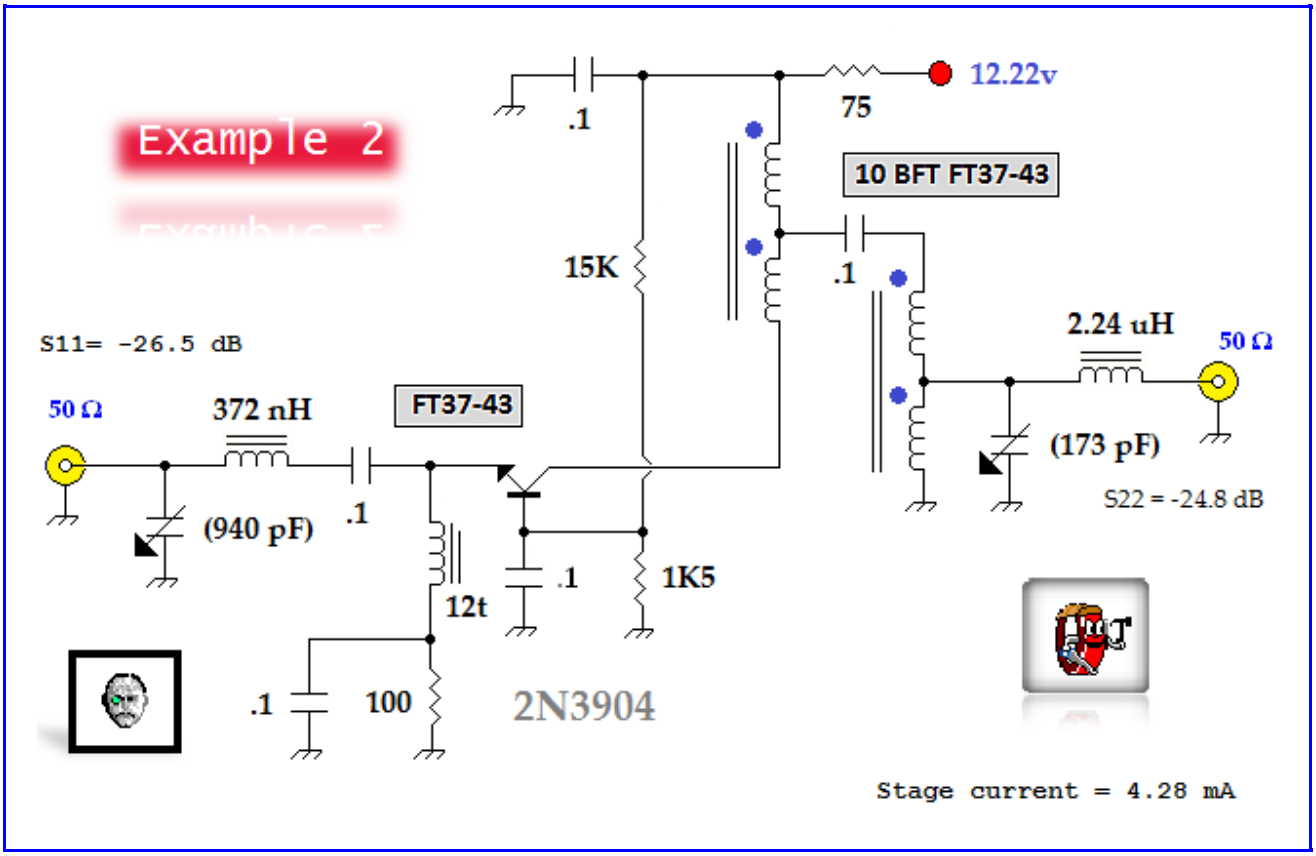
Above — AC voltage with reverse gain set-up: 5.28 mV pk-pk.

S21 = 15.1 dB. S12 = -31 dB. I tested my driven amp's linearity by adding a scrap, standard value 6 dB pad in between the signal generator and the amplifier — the S21 was 15.2 dB with the pad and 15.1 dB without the added 6 dB pad. It's linear.

Reverse Isolation Calculation

R.I. = S21 - S12
R.I. = 15.1 dB - -31.03 dB
R.I. = 46.13 dB

Example 2: Common Base Amplifier



Above — A common base amp employing L- networks for a strong return loss in and out at 7.039 MHz. S11 and S22 = the negative of return loss.

Through connector voltage = 192 mV peak-peak.

S21 = (1.88 volts peak-peak AC voltage) = 19.82 dB.

S12 = (1.68 mV peak-peak AC voltage) = -41.16 dB.

Reverse isolation at 7.039 MHz = (S21 - S12) = 61.42 dB

I confirmed the linearity of my S21 using the aforementioned 6 dB attenuator pad — I couldn't increase my signal generator output level above the indicated -10.35 dBm, since gain compression emerged in the common base amplifier (reduced AC voltage was measured).

My reverse AC voltage was under 2 mV; about the threshold where my oscilloscope waveform becomes rather ugly and sits in the noise. Clearly, the limitations of measuring reverse isolation with an oscilloscope must be factored. Still, you got to love 60 dB + of reverse isolation in a popcorn circuit.

1 local professional EE told me if you can measure it with a 'scope, you don't have spectacular reverse isolation - no doubt, a spectrum analyzer

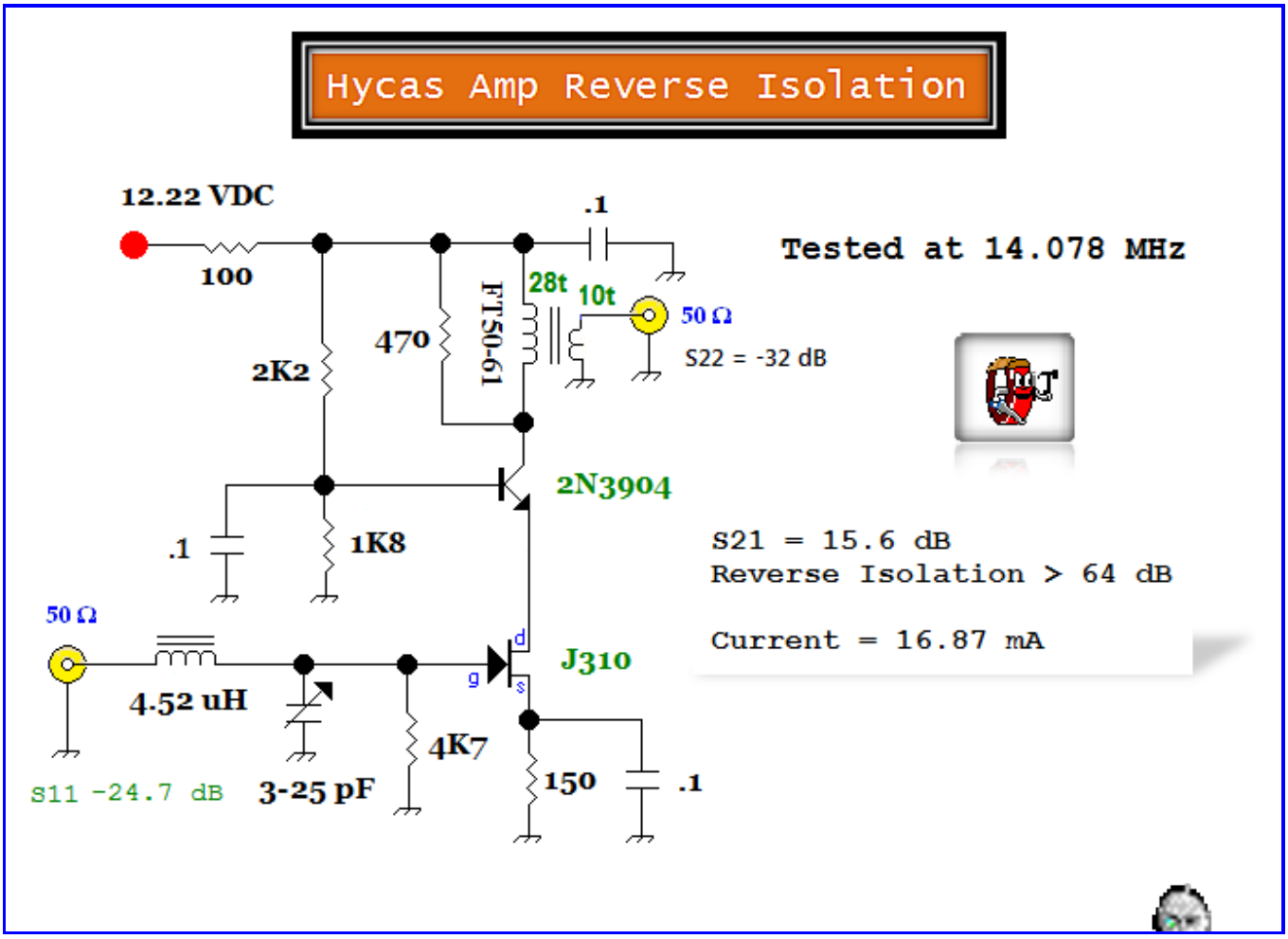
pumps up the measurement quality in circuits with high reverse isolation, but more amateur designers have 'scopes than spectrum analyzers, so just do your best.

Involving Scattering Parameters as possible on your workbench can only lead to circuit improvement. You can't better your outcomes if you can't or don't measure them — applying and more importantly, understanding test equipment is just 1 component of our hobby. A hobby unto itself; test equipment activity complements amateur radio design. I have met test equipment focused builders who make radio gear just as an excuse to apply their test gear!

Improved bench practices are the corollary of striving to learn more about measurement techniques and increasing our collection of measurement devices.

QRP — Posdata 1: Hycas Amplifier

I love the hybrid cascode (hycas) as a general purpose RF amplifier. What's not to love about a using a common source FET followed by a common base bipolar amp? I attempted to measure the reverse isolation of a version just using an oscilloscope.



Above — A hycas amp set up for high return loss on both ends at 14.078 MHz. Too much current may cause gain compression and harmonic distortion, so please test your hycas amps for both. I tested using a signal generator with a 30 dB return loss driving a 50 Ω terminated oscilloscope. Since the hycas amp contain a high impedance input JFET and a common base amp, the reverse isolation should be reasonably high, or at least as good as a common base amplifier.

My testing failed — The reverse isolation was too high to measure with an oscilloscope. Using proper bench techniques (linear amplification + honest scope reporting), I determined the highest reverse isolation I could measure = 64 dB. In fact, injecting a whopping signal of 1.08 volts peak-peak into the hycas output port only gave an S12 of 1.84 mV — whoa!

The problem is such a strong signal (1.08 volts peak-peak) at the input port results in severe limiting and distortion; so valid reverse isolation measurement isn't possible. Even a 350 mV peak-peak signal may give some gain compression during S21 measurement depending on your matching. Thus, I can only accurately say that the reverse isolation of my hycas amp is greater than 64 dB.

Strong reverse isolation is 1 reason I favor hycas amps as VFO buffers. They make pretty good I.F. amps also.

The hycas IF amp system by Wes, W7ZOI and Jeff, WA7MLH offers amazing performance and features an excellent JFET bias scheme. I built 1 in 2008 — amazing design.

QRP — Posdata 2: Doesn't S12 = Reverse Isolation?

Many web sites, books and people report that reverse isolation = S12, yet above, I depart from this argument. In truth, I think reverse isolation equals S12, but reverse isolation may also equal S21 - S12.

I'll let you decide what to do, but explain why I enjoy the latter.

S12 is a negative value.

I prefer to turn that negative value into a positive 1 — RF Workbench 4 concerns measuring and applying amplifiers with the goal of high reverse isolation and not just measuring S12.

The main purpose to quantify amplifer reverse isolation is to strive to *improve reverse isolation* and an amplifier is but 1 component in our 50 Ω block. I believe in a creative, systems approach; open minded and positive (pun intended).

The whole RF Workbench series attempts to present 50 Ω bench measurements in a vibrant way devoid of excessive + boring engineer-style content that could blank the eyes of the budding Hams/SWLs designers that visit my site. I imagine this web site bores more advanced RF designers *to tears*.

Our goal is to obtain high reverse isolation while applying a 50 Ω systems approach.

Break away from strict + "stodgy" math-driven methods to fuel creative thought and experimentation.

I posit that an appropriate figure of merit in a well designed isolation amplifier is the difference between S21 and S12; and therefore, the term reverse isolation can be more than just S12.

Ham/SWL component-level experimentation by commoners like me is ~~slowly~~ dying and being replaced by a new generation of skilled, code writing experimenters. Although, some builders just copy other people's code and then apply it to kitted hardware. Reverse isolation impacts both our analog and digital designs.

Please consider this example:

I'm switching a level 7 diode ring mixer with 12 MHz and per normal, create lots of internal harmonic energy. With my spectrum analyzer connected to the mixer RF port I measure my 12 MHz LO signal at 50 dB below a +7 dBm signal, or -43 dBm. That's a 50 over S9 signal — very high amplitude in context!

I require strong LO isolation in my circuit and thus stick in a 50 Ω input/output amplifier. I measure this amp: S21 = 15 dB and S12 = -31 dB.

So, the signal at the amplifier input is -43 dBm plus -31 dB = -74 dBm. But, alas, -74 dBm isn't good enough me — I want to use that amplifier to elicit greater isolation. However, I don't want any gain in my system, so I insert a 15 dB attenuator pad after the amplifier. For this pad, both S12 and S21 = -15 dB. For my amp, the net cascade is S21 = 0 and S12 = -46 dB.
Since S21 = 0, the block has 0 impact on the signal amplitude applied to the mixer, but the signal at the input of my isolation amplifier is -43 dBm plus - 46 dB, or -89 dBm. This isolation I like — it also illustrates a systems approach that gets you thinking about measurement in your own 50 Ω blocks.

The figure of merit for making a good isolation amplifier is now the difference between S21 and S12. If you want, go ahead and just use S12 for reverse isolation, but you'll probably measure S21 plus S12 anyway and that's what this web page is about! Onward.

QRP — Posdata 3

Comments From the Workbench

I'm no amateur electonics expert — I'd like to be one, but this is a tough field; RF and AF *design* is quite scientific, under-resourced and a bit overwhelming. How do we experimenters advance and stay motivated? Reading works by professionals like Chris Trask, N7ZWY, Bob Larkin, W7PUA, Doug Self, Rod Elliott and others may highlight our lack of knowledge and scientific methodology — a realization which can distress and demotivate us lay-designers. To a degree, this is irrational thinking; personal growth is always about hard work, problem solving and overcoming barriers.

Unlike the white belted Karate student, who studies and practices under the guidance of a master to attain black belt skill level, most amateur designers, excluding electrical engineering students, can't access good teachers. As a lay-person, with few face-to-face mentors (nobody in

Canada), I try to learn by experimenting and incorporating whatever knowledge, advice or schematics I can find. Fortunately, some Electrical Engineers give me advice by email and in turn I'm able to share this information via experiments on QRP / SWL HomeBuilder.

Our dusty, analog hobby fades palpably — the number of analog electronics gurus dwindles each decade and modern electronics embraces miniature circuitry often involving digital ICs controlled by lines of code.

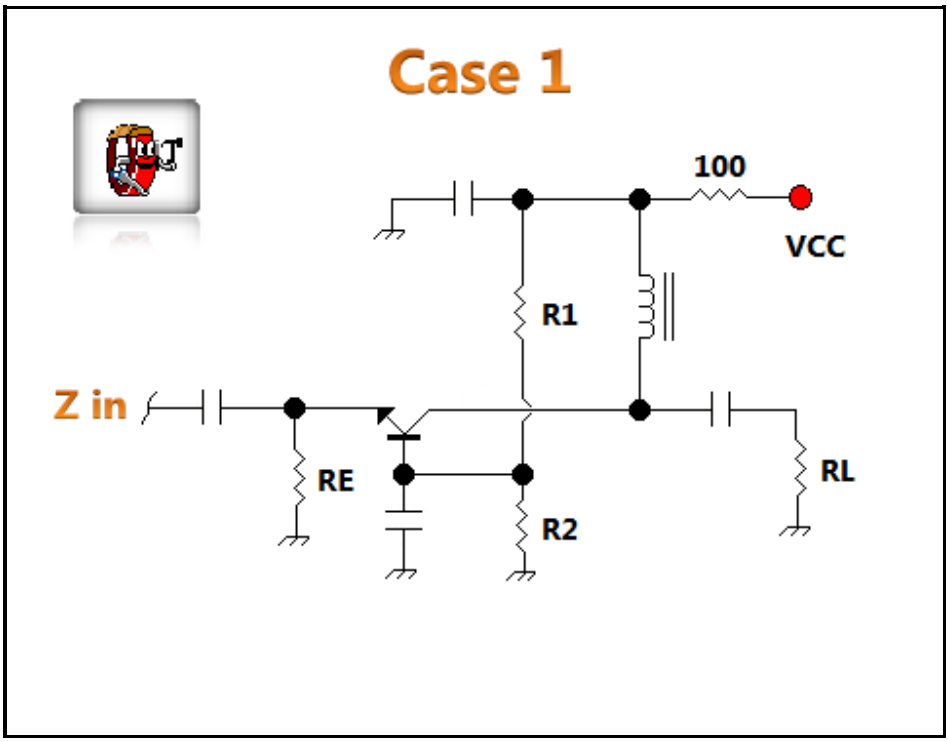
Current electronics hobbyist magazines rightfully focus on topics that are contemporary or important to their advertisers; for example, promoting mixed-signal ICs, DSP, microcontrollers and the kits they describe and then sell for income. Nuts and Volts is 1 example. Both analog RF and AF design increasingly lies in the hands of a small group of specialists, enthusiasts and students.

Yet, we *persevere*. Sharing our knowledge, circuits, experiences and references on the Internet helps sustain our small global community. That's the site purpose— sharing the (warts and all) experiments + basic information of a lay-person.

The Emitter Choke in Common Base RF Amps

This web page covers reverse isolation — a really important topic. 2 principle amplifiers we employ for strong reverse isolation are the common base BJT and common gate JFET alone or in cascode with other amplifier topologies.

Some comments regarding using a radio frequency choke in the common base amplifier follow.

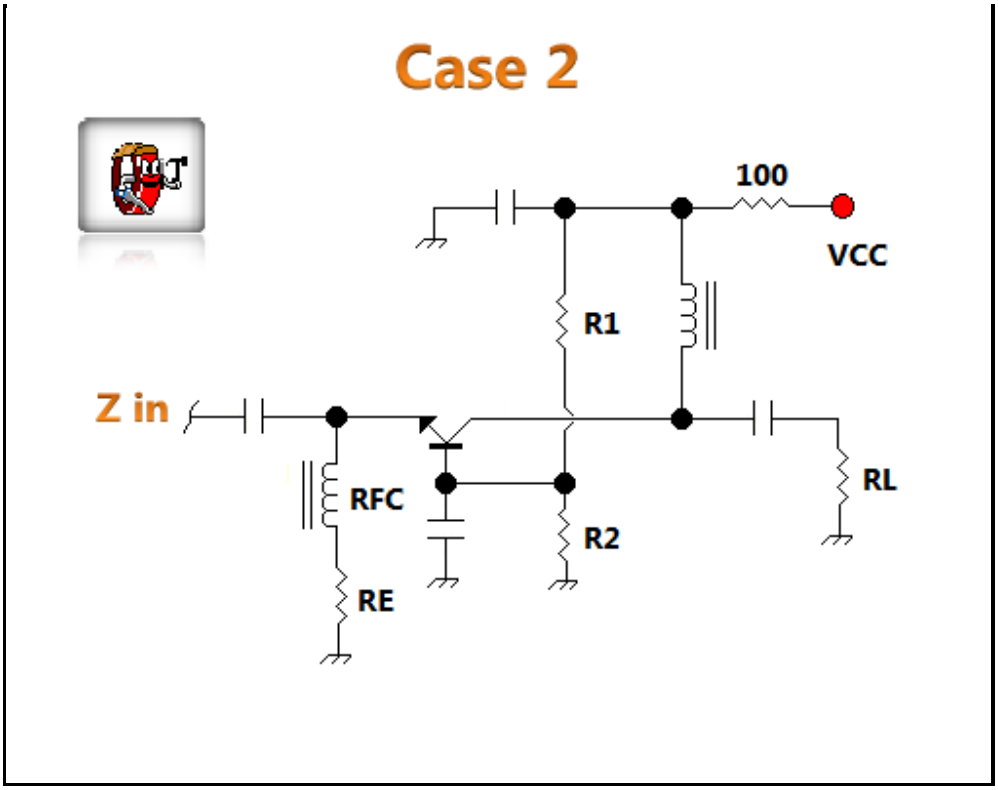


Above — Case 1: Emitter resistor only.

Apart from providing DC bias along with R1 and R2, emitter resistor RE plays another important role. Despending on its value, a portion of the input AC signal may pass through RE to ground instead of going through the transistor — degrading signal amplitude and noise figure. To minimize this, the resistor value should be many times (~10X or more) than the input impedance of the amplifier.

Although we might bias a common base amp to give an input Z of 50 Ω, often we'll choose a much lower input Z to get higher voltage gain. Input Z = 26 / Ie where Ie = mA; so if you bias for 5 mA, you are looking at an input Z of ~5 Ohms. In that case, a low value bias resistor such as 100 Ω won't shunt much of the input signal to ground, nor will it likely contribute much noise.

For most common base RF amps, a correctly chosen emitter resistor is all that's needed to decouple the AC signal and using an emitter choke proves hard to justify. However, it's important to understand how to apply an emitter choke since the basic principle also extends to the common gate JFET amplifer and other circuits.



Above — Case 2: Emitter resistor plus a choke.

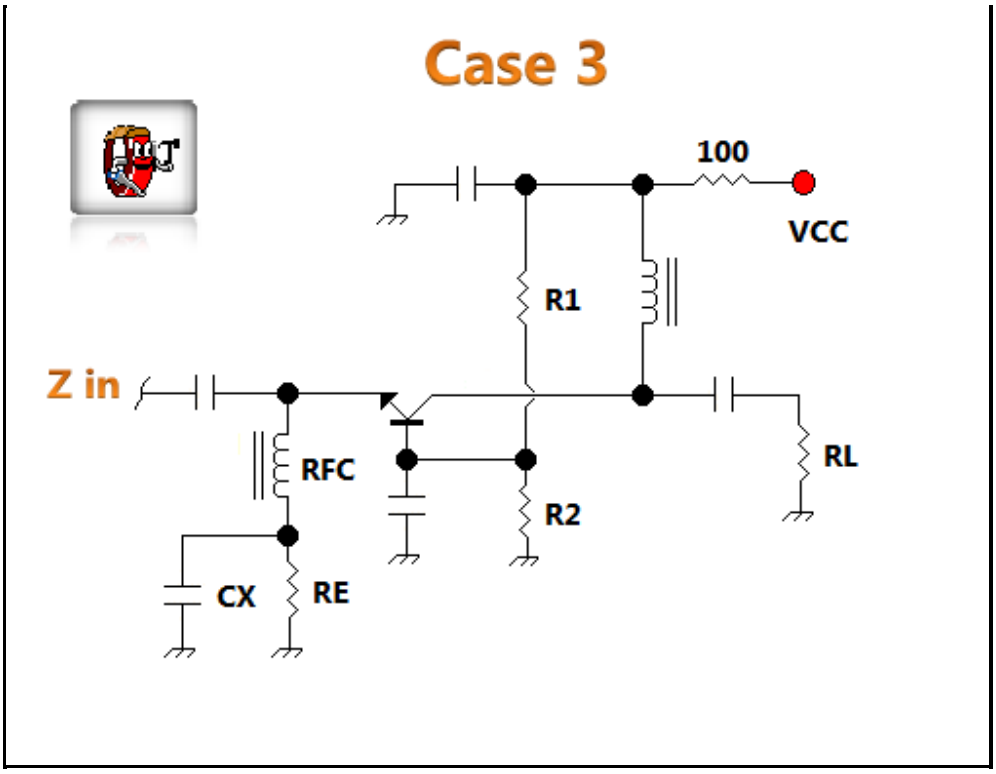
The choke's main purpose is to block or choke RF from passing to ground. The ideal choke would present infinite impedance to AC signals, plus 0 resistance to DC voltage. In reality, "ideal" = fantasy electronics and you can simply estimate a choke's inductive reactance using the classic formula (XL =2*PI * Freq * L).

Using a coil (and not just a resistor) is generally better for decoupling — although how much better might be debatable. If the inductive reactance (XL) of the coil is significantly higher than the input impedance of the transistor, then all of the input signal power goes to the transistor.

By convention, a minimum choke XL should be at least 3 times the input resistance, however, the self-resonant frequency of the coil must be significantly higher than the applied frequency. Thus, an ideal range of inductive reactance exists, and too little or too much can degrade performance. Many builders target an XL around 10 times higher than the transistor input impedance at the lowest operating frequency.

Example: For a common base amp biased for an input Z of 50 ohms, the minimum inductive reactance (XL) for the choke = 500 ohms. To calculate the inductance of an emitter choke for this amp at 50 MHz, we re-arrange the formula to solve for L.

L = XL / 2* Pi * F
L minimum = 1.59 uH.
Winding the choke on a ferrite core, or possibly a bead for VHF often means less turns, less winding capacitance and a higher self-resonant frequency.



Above — Case 3: Bypassed emitter resistor plus choke.

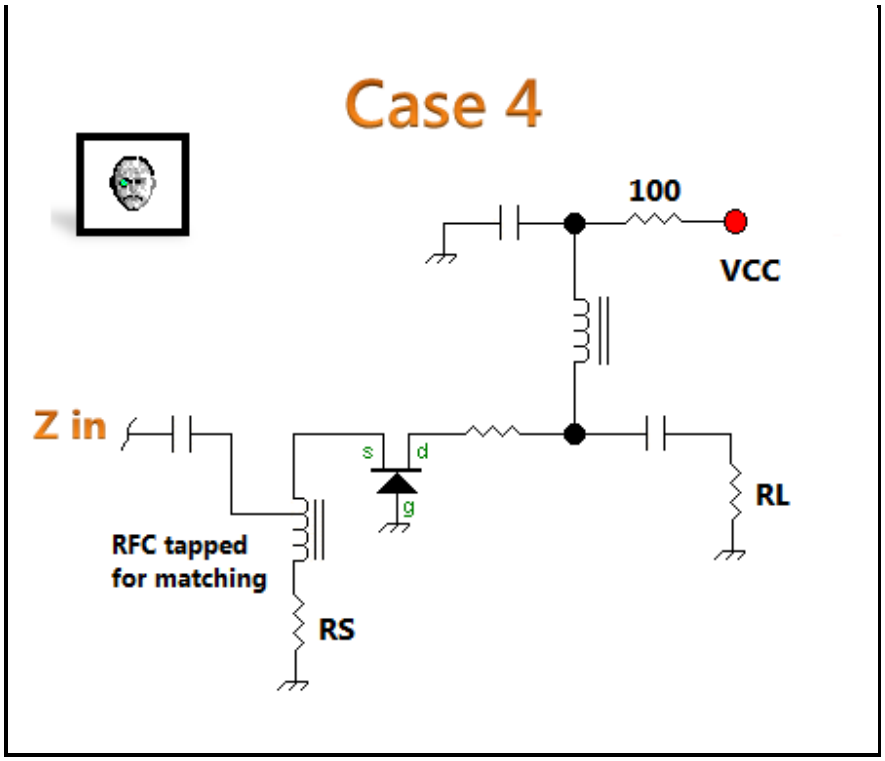
The primary purpose of the capacitor across RE is to filter resistor noise — but that is only an issue well below the frequency of interest and it should not be relevant at high frequencies where the choke reactance is significant. There may be some useful effect if the self-resonant frequency of the capacitor Cx is above the frequency of interest.

You can only use bypass capacitor **Cx** when a choke is implemented.

A 0.1 uF may be useless at high frequency. In error, I've used this value previously on the site; after 14 years of experimenting, I've learned a lot from my design mistakes.

The case of the common gate JFET amplifier

This discussion also informs common gate JFET amplifer design. The JFET source requires signal decoupling similar to the emitter of the bipolar transistor discussed above.

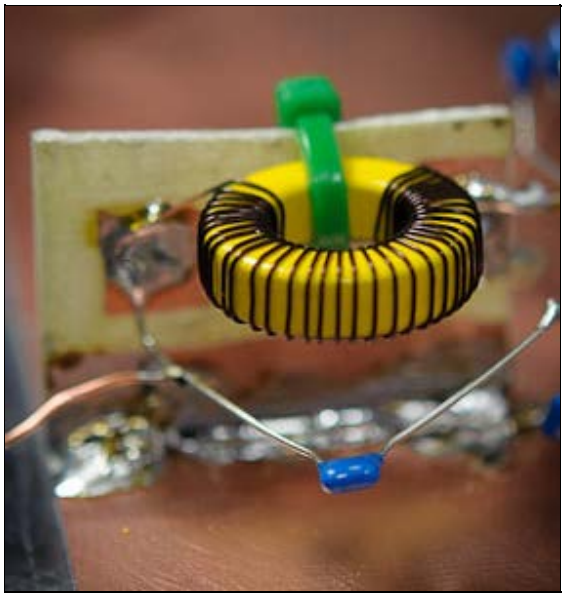


Above — Case 4: A choke plus source resistor will commonly be the "go to" design. Things get a bit more complicated with some, but not all JFET circuits — engineers often match the JFET input for a low noise figure rather than just the "correct" input impedance.

A good example follows: We might place a common gate JFET amp after a diode ring mixer because of the wideband load it presents to the mixer's RF port. The best noise match may occur with a hypothetical input Z of ~70 Ω (this argument represents an advanced topic).

After measuring the JFET pinchoff voltage and Idss, you would likely find that a source bias resistor of ~100 Ohms would be needed. This R value is so close to the JFET input Z that signal losses to ground would occur — demanding a choke for signal amplitude preservation, plus impedance and noise figure control.

VFO - 2011



Building VFOs in 2011 might seem an irrelevant exercise given the move to and evolution of digital signal generators laden with bells and whistles like memories and audio or video frequency displays.

A successful L-C VFO requires skill, patience and some good parts to pull off — else, a "drift monster" may result. Despite their limitations, it's possible to build L-C VFOs with low frequency drift, distortion and phase noise; our typical VFO performance markers. L-C VFOs don't require programming skills or equipment to encode a microprocessor — making them a good choice for people who don't build or can't afford kit oscillators. Most of all, they kindle creativity, problem solving and pride when your oscillator actually works as planned. Junk box radio; my passion.

This material reflects lots of empiricism; not pure science. It's really *your* VFO design odyssey; a chance to think creatively and critically to sort out what works and what's folly.
Countless web pages discuss VFO design and I encourage you to search for and read them. Wes' EMRFD oscillator and temperature compensation notes = essential reading. Only your first 25 VFOs will prove difficult — it gets easier after that.

VFO 2011 Topics:

1. Frequency Stability Notes
2. Vackar VFO Experiments
3. HF Signal Generator
4. Miscellaneous Bits

1. Frequency Stability Notes

Building an oscillator that stays on frequency purports our greatest challenge and goal in L-C VFO design. Since drifting VFOs pose a source of frustration, I cover some topics that may help your VFO stay on frequency — do they help?

What is good drift parameter?

I'm uncertain, for after warm-up, I've measured kits that drifted 50-150 Hertz per hour, built L-C VFOs that drifted under 20 Hz per hour and every once and a while, build a drift monster VFO that sweeps upward at 2 - 8 hertz per minute! Likely under 20 Hertz per hour after warm up = a gold standard to compare against. You should be able to listen to a 10-20 minute QSO with no re-tuning, however, this assumes the transmitting stations are locked on frequency.

1. Unloaded Q and Frequency Stability

The number 1 reason to employ high resonator Q in oscillators is to obtain low phase noise. Secondly, the very steep phase slope through high Q resonance minimizes the effect of amplifier phase shifts caused by temperature changes and this in turn, minimizes any amplifier-induced frequency instability.

Long term frequency stability is chiefly dependent on the temperature, environmental and age stability of the resonator components regardless of Q.

I often see designs featuring high Q inductors wound on powdered iron toroids complimented with trashy, low Q variable and/or fixed capacitors, If you design for a high Q tank to minimize phase noise, consider using a high Q coil plus appropriately temperature stable, high Q capacitors.

2. Temperature Stable Inductors

Knowing that I'm venturing into a topic of great debate and lore, the inductor is 1/2 of the VFO resonator and thus a major source of temperature drift in L-C VFOs. Since MF and some HF VFO designs *may* preclude using the *inherently more temperature stable air wound inductor*, powdered iron toroids dominate our evermore compact designs. Many builders choose #6 material, although the lower temperature coefficient of #7 material theoretically should be better — however, my experiments have failed to measure a significant difference between these 2.

Some builders prefer size 68 inductors, for the bigger core is less affected by heating than smaller size toroids. My experience suggests that providing the VFO amplifier current is kept low, both size 50 and 68 are both suitable and the inductance needed should inform the core size.

I used to think that heavier gauge wire created greater frequency stability than smaller gauge wire until Wes, W7ZOI, woke me up. As it turns out, smaller gauge wire is often better for thermal stability because smaller gauge wire lies closer against the toroid core. Winding stiffer, heavier gauge wire creates more air gaps than smaller gauge wire and air gaps expand and contract during temperature changes. Smaller gauge wire will have a reduced Q, but it won't be as significantly lower as you might guess. As possible, I prefer tightly wound number 28 wire. 26 gauge wire tends to be my maximum size wire for VFO coils, however I suggest you make your own conclusions.

Wash your hands before winding and use both hands to actively move both the toroid and wire for tight turns. Take your time, ensure steady wire pressure and avoid kinking your wire. Taps increase the likelihood for air gaps — mitigate this by stripping the 2 tap forming wires as close to the toroid as possible and twist them into 1 wire right down tightly to the toroid edge to reduce any air gap.

The thermal stability characteristics of wire can be mitigated somewhat by annealing the wire with temperature cycling or by dunking it in boiling water. Roy, W7EL first reported annealing coils in 1980 and this has been confirmed during experiments by builders using temperature controlled ovens. I don't boil my coils any more.

3. Double Stacked Toroids

I noticed a new trend in VFO design is to stack 2 powdered iron toroid inductors. This allows the builder to double the inductance per number of windings over a single toroidal inductor. In an L-C VFO, the goal of these builders possibly is to reduce heating effects, increase unloaded Q, or perhaps to reduce core magnet flux density. For me the goal is far simpler, I just want to make compact, large L value inductors for 3 MHz and less.

Above — A T68-6 hamburger. The two T68-6 cores were epoxy glued together and compressed lightly in a vice for several hours. One of the initial tests I performed was to see if boiling the stacked coil affected the epoxy glue. The glue was not effected by annealing wire on a stacked coil with 5 or even 10 minutes of boiling in water. As mentioned, I stopped boiling my VFO inductors as tightly winding them with 26 gauge wire seems to work well.

I hold concern that stacked toroids may create more wire-air gaps when compared to a single toroid and stay with 1 toroid as possible. In compact antenna tuners and other non VFO projects, this isn't an issue.

4. VFO Tank Capacitors

We choose VFO tank capacitors to avoid temperature change caused frequency drift, or to counter drift during our temperature compensation process.

Many authors have published guidelines for long term temperature stability. It's important to consider these guidelines, but also try whatever works. I believe the following arguments are accurate based upon my experiments:

1. Multiple NP0 or C0G (0 temperature-compensation) tank caps: Most builders minimally use 4 or more C0G or NPO capacitors to reduce heating effects and to average out temperature coefficient variations.

2. No VFO tank capacitors from online surplus parts stores; buy new stock from known and reputable manufacturers. Grab bags and musty, old, surplus parts can obviate good design.

3. Trimmer and tuning caps need to be temperature stable. Air variable capacitors = my favorite, as possible.

4. Varactor, or diode tuning generally = more drift and a greater need for temperature compensation.

5. Employ short, stiff capacitor leads. I use 100 volt or higher voltage C0G tank caps as they tend to have thicker leads that stay put — perfect for Ugly, Manhattan, or Chuck Adam's MUPPET construction.

5. Temperature Compensation

The goal of temperature compensation is to cancel the tendency of the VFO to drift in 1 direction — easier said than done + very time consuming. A web search for VFO temperature compensation will yield many good write-ups. I feel it's partly art, partly luck and partly science. Your net VFO temperature coefficient can be affected by so many variables, so no 1 recipe will ensure a low drift VFO. Experiment, allow a lot of time to assess your changes and be patient — you'll figure it out.

The simplest way to test for drift involves watching a frequency counter, but if you don't have one, you might use a commercial, frequency stable (synthesized) receiver set in the SSB/CW mode. I use both. Experienced builders often employ an oven to test their temperature compensation at different, controlled temperatures. Wes, W7ZOI employs a styrofoam cooler housing a light bulb heat source controlled by a Variac. See EMRFD for more details and a photograph.

In 2011, I decided to build up a supply of temperature compensation capacitors and keep them in their own parts bin.

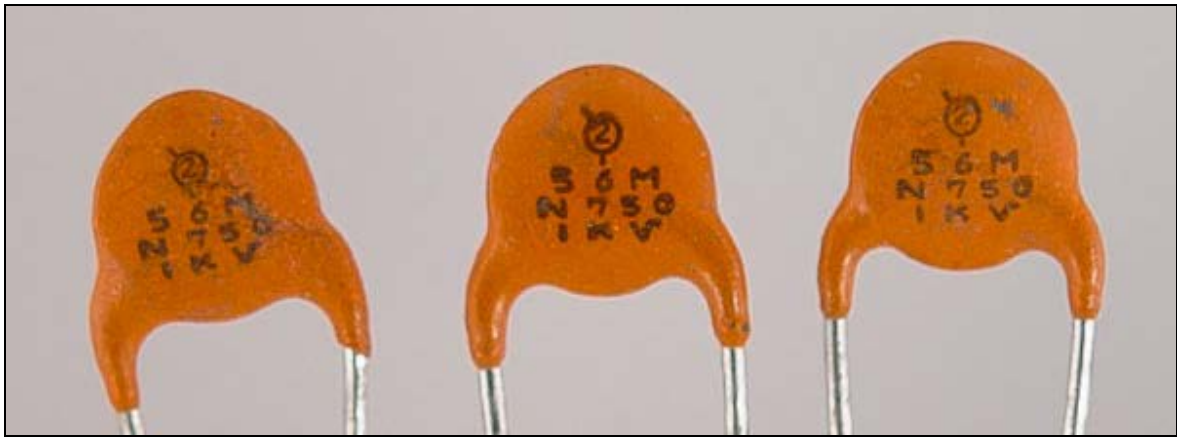
Above — "Tempco caps". A parts drawer containing polystyrene capacitors from 10 to 270 pF plus some 56 pF ceramic N750 capacitors for negative temperature compensation. I purchased these capacitors on eBay.

For capacitors other than NP0 (which use 0 instead of a ppm value), the temperature coefficient = **P** for positive and **N** for negative, followed by a 3-digit value specifying ppm/°C. For example, N220 is - 200 ppm/°C. and P100 is +100 ppm/°C.

I use NP0 and C0G ceramic capacitors interchangeably for both tuning and RF bypassing the VFO tank resonator. For C0G/NP0 *temperature compensation bypass*, I normally apply 0.01 or 0.001 µF caps, however, the more expensive 0.1 µF COG ceramic capacitors are still sold if you need C0G/NP0 bypass <= 7 MHz.

If your VFO is drifting upward you might insert 1 or more positive coefficient capacitor(s). If your VFO drifts downward, then try using negative coefficient value(s). Sometimes just 1 capacitor will do the job.

Since I don't stock any positive coefficient capacitors for positive coefficient compensation, I might try a adding a silver mica capacitor. ***Caution*** silver mica capacitors are extremely non-predictable and can't be universally recommended in temperature compensation schemes. You might also try swapping out 1 or more of your main tank NP0 or C0G capacitors in case they are bad; sometimes it gets frustrating. I provide some temperature compensation examples on the QRP Modules 2011 web page in the 7 MHz VCO section.



Above — 56 pF N750 ceramic capacitors rated at 1KV

6. Mechanical Rigidity

Movement of your VFO tank parts may lead to frequency instability. For example,

1. Well secure your single-sided only copper board. I use at least 4 number 8 bolts — 4-40 hardware is too light. Boards can warp over time if not lashed down properly. Aggressively bolt down any variable capacitors. No tank parts should move.

2. Anchor your inductor so it cannot budge: nylon bolts, zap-straps, glue - whatever.

3. Consider placing the VFO in a strong chassis with rubber feet.

4. Buss wires should be made from thicker gauge, well anchored wire.

7. Miscellaneous Points

1. Regulate the VFO amplifier DC voltage and wideband filter it. Voltage regulators require RF and often AF bypass to attenuate any noise or ripple riding on the DC.

> A decoupling resistor with a bypass cap on either side will widen your DC supply filtering bandwidth and deserves strong consideration. A poorly filtered DC supply can easily transmit the VFO tank energy to other stages along your DC lines and also may allow noise on the DC supply to modulate your VFO and increase phase noise.

2. You should have the buffer + a load resistor connected to your VFO when testing. **Do your temperature stability work after the buffer is built and the VFO is in its case.**

3.Stick your VFOs in an air tight, RF tight case to minimize air temperature changes and RF leakage respectively. Sometimes a VFO will drift once in a case because any radiated buffer amplifier heat will warm up the inside of the chassis. This usually levels off after warm up.

4. Modern voltage regulators may significantly reduce noise compared to a zener diode regulator. Specific low noise and low temperature coefficient voltage regulators are available, but maybe overkill for you. Whatever you use — filter it well —
The Micrel MIC5209-5.0BS in SOT223 sits in a couple of my reference oscillators.

5. JFET Gate clamping diodes may increase phase noise, but not prohibitively so in most popcorn designs.

6. When winding toroid inductors, wind 2 extra turns. When finished, unwind the first 2 turns since they are usually loosely wound and prime culprits for air gaps.

7. Since magnet wire comes off small spools, wire has a natural curve or radius — ensure you wind your coils according to the natural curve of the wire.

8. The need to secure powdered iron windings with dope, wax, goop, etc. is over-emphasized and usually unnecessary.

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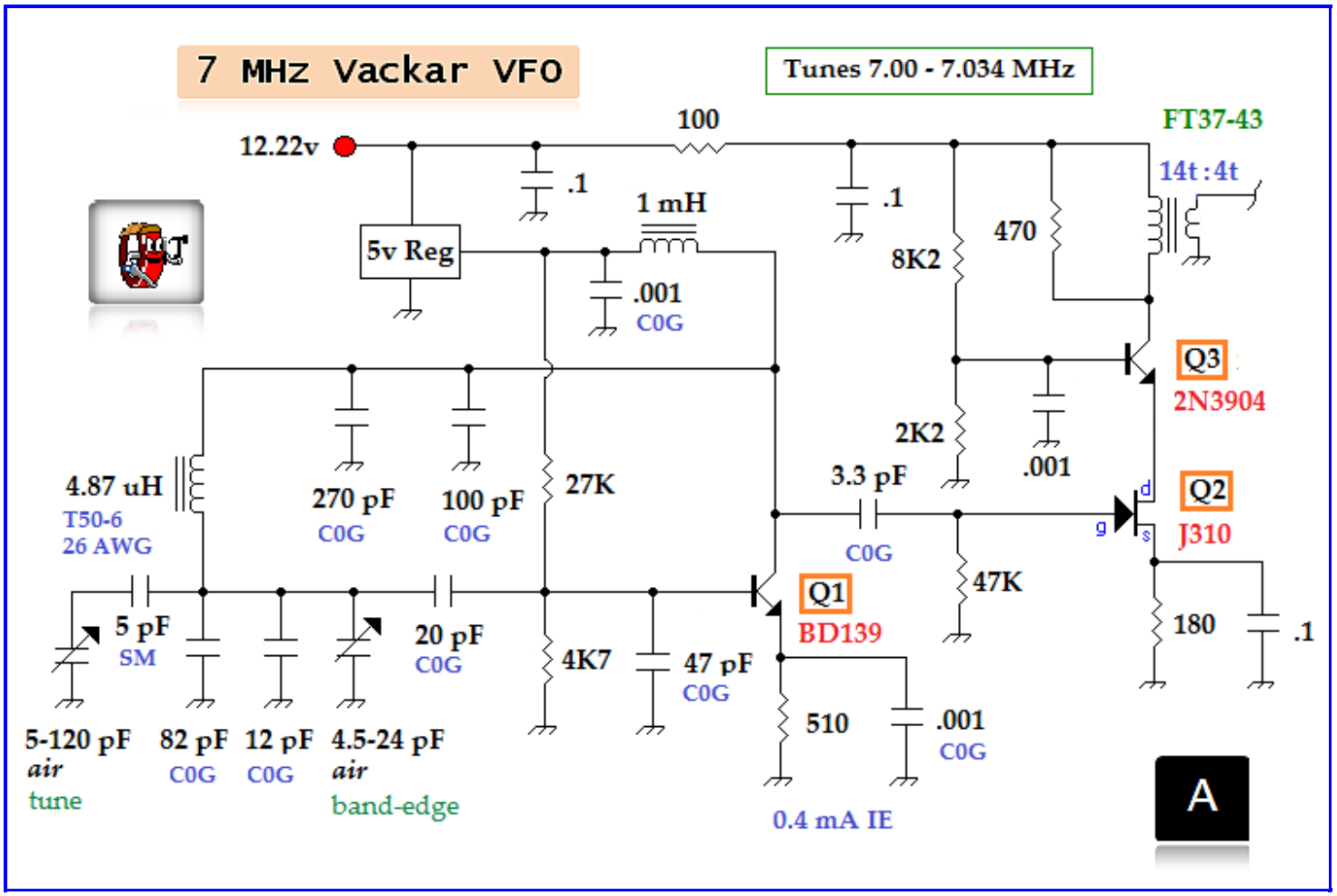
2. Vackar VFO Experiments

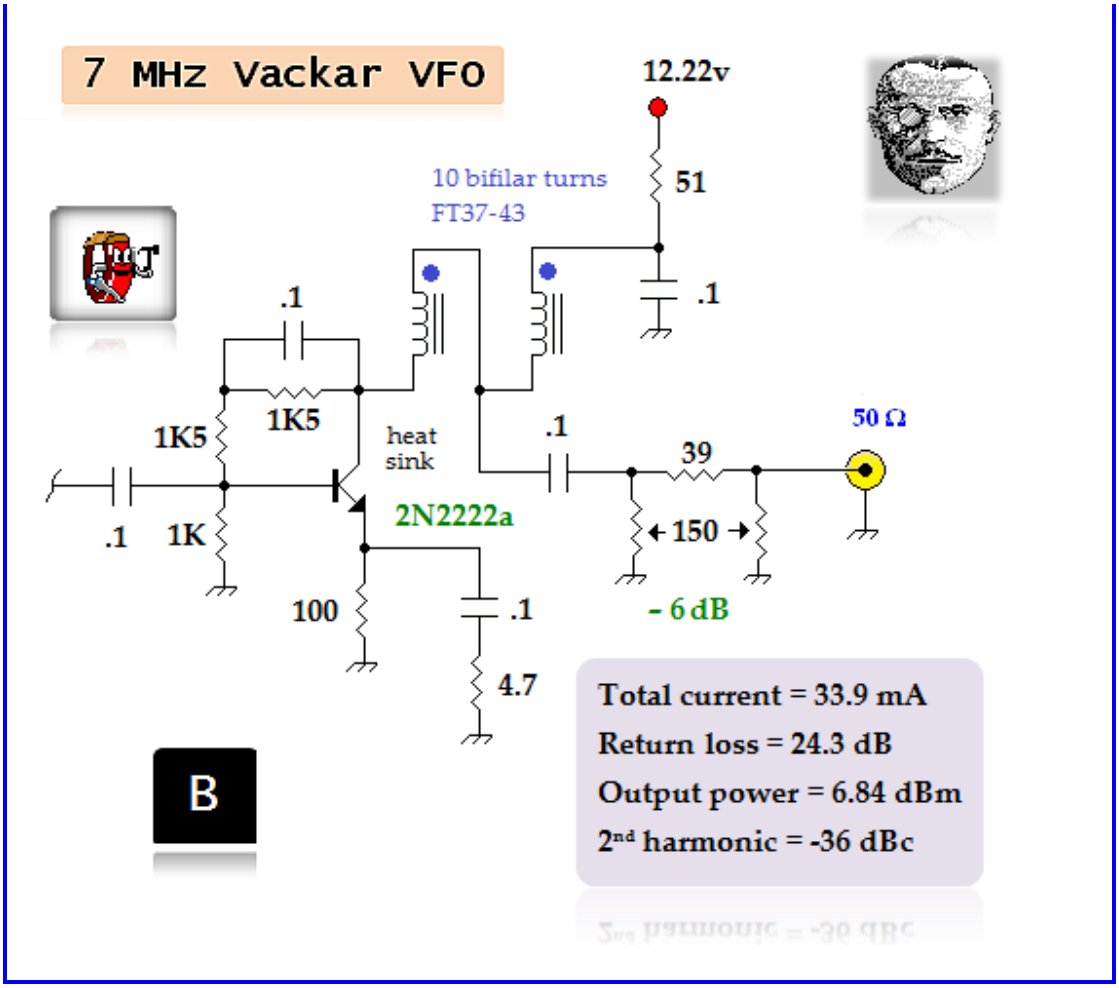
Some builders proclaim the Vackar as the "King of VFOs". I built a couple and became impressed by the low distortion and less than 5 Hertz per hour long-term drift achieved in my 2 designs. Inspired by work from Iulian, YO3DAC entitled Very Low Phase Noise Vackar VFO for HF Transceivers *(link and reference used by permission of Iulian)*, I crafted my version from his notes and schematic.



Above — Schematic of the Vackar VFO employing a BD139 — a large area transistor, to reduce 1/f noise. Iulian shared many design pearls in his paper and I won't repeat them. I ran Q1 with 0.4 mA emitter current to reduce heating and flicker noise. It's difficult to measure flicker noise, so no objective comments can be made.

I limited the tuning range to 34 KHz since the tuning capacitor lacked reduction gear and I was born with fumble fingers. As a CW operator - you'll find me down at the bottom of the band away from the RRTY anyhow. Increasing the 5 pF cap coupling the tuning capacitor to the tank increases the tuning range as expected.

Temperature compensating my VFO with the 5 pF silver mica capacitor proved a gamble since SM caps are unpredictable and often best avoided. In *my* VFO, it worked perfectly, however. This circuit is difficult to replicate and *not recommended* for new builders. Temperature compensation provided the sublime frequency stability.



Above — The final amp and measured output data. Measuring the 2nd harmonic 36 dB down without any tuned circuit or low-pass filter rocked. I ran nearly 22 mA of emitter current to bump up the return loss and spectral purity. A 2N5109 or 2N3866 would likely do a better job with less current. Total current = the entire VFO current. I glued a drilled copper penny on the 2N2222 to dissipate heat.

A photo of my version of a Vackar VFO. My design goals included low phase noise, low distortion, a return loss over 20 dB, good reverse isolation and ~7 dBm output power. I believe all VFOs are experimental; you build to suit whatever tuning capacitor or varactors you have, plus design around constraints such as total current, tuning range and other personal criteria.

Unlike harmonic distortion, oscillator phase noise, being close to the oscillation frequency, cannot be removed by filtering nor limiting — you must design for low phase noise. Modern digital VFOs are well harmonically filtered, and any phase noise depends on the DDS clock employed, so check the DDS specifications carefully if you go the DDS VFO route.

I'll be the first to state I'm no expert with VFOs, however, likely the only way to become expert is to build many and learn from your mistakes.

Above — A 7 MHz Vackar VFO with the lid off

Sound Test?

Although this technique raises the ire of some builders, I test my VFOs in a nearby receiver. The VFO output was terminated with a 51 ohm resistor that was also attached to my frequency counter via alligator clips and wire. I tuned a nearby CW superheterodyne receiver to 7.00 MHz with the audio beat note centered in its 600 hertz wide I.F. filter and watched the counter plus listened to the receiver.

Click for a 1 minute 32 second audio file of the result (it stayed perfectly on frequency for ~5 hours of testing before I got a bad headache from listening to it and shut it off). *You can initially hear a station in the back ground despite only having a 45 cm piece of wire as the receiver antenna.* The VFO slowly drifts down to 6999996 Hz and then slowly back up to 7000000 Hz. You can hear the signal amplitude decrease as the VFO drifts down. So it doesn't stay perfectly on frequency, but slowly cycles up and down a few hertz. This VFO is my lab temperature stability benchmark for an L-C VFO.

A badly drifting VFO will move out of the test receiver I.F. pass band and sound like a Theremin as it does. Testing in a receiver; places the VFO in the exact circumstance it will be used — beating RF to mix to another frequency; in this case, base band audio.

The temperature stability and compensation of any VFO schematic are rarely reproducible since there are just too many variables. Try your best to get the drift out of your VFO using low temperature coefficient capacitors (NP0/C0G) and then after that, temperature compensate. Even today, I occasionally build a drift monster VFO and become frustrated. VFO design is not for the faint of heart and it's no wonder that many builders make a VXO, or cave in and build or buy a DDS signal generator.

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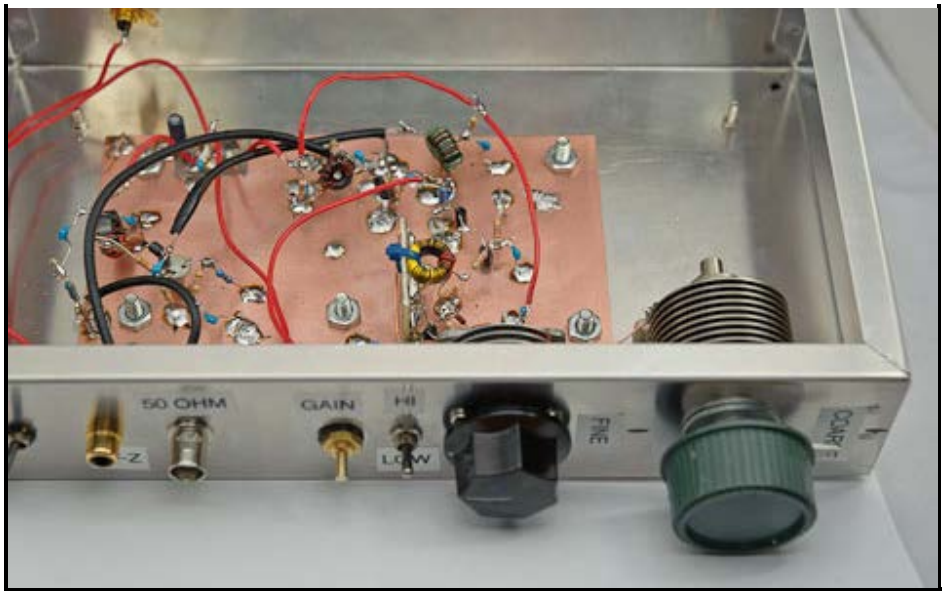
3. HF Signal Generator



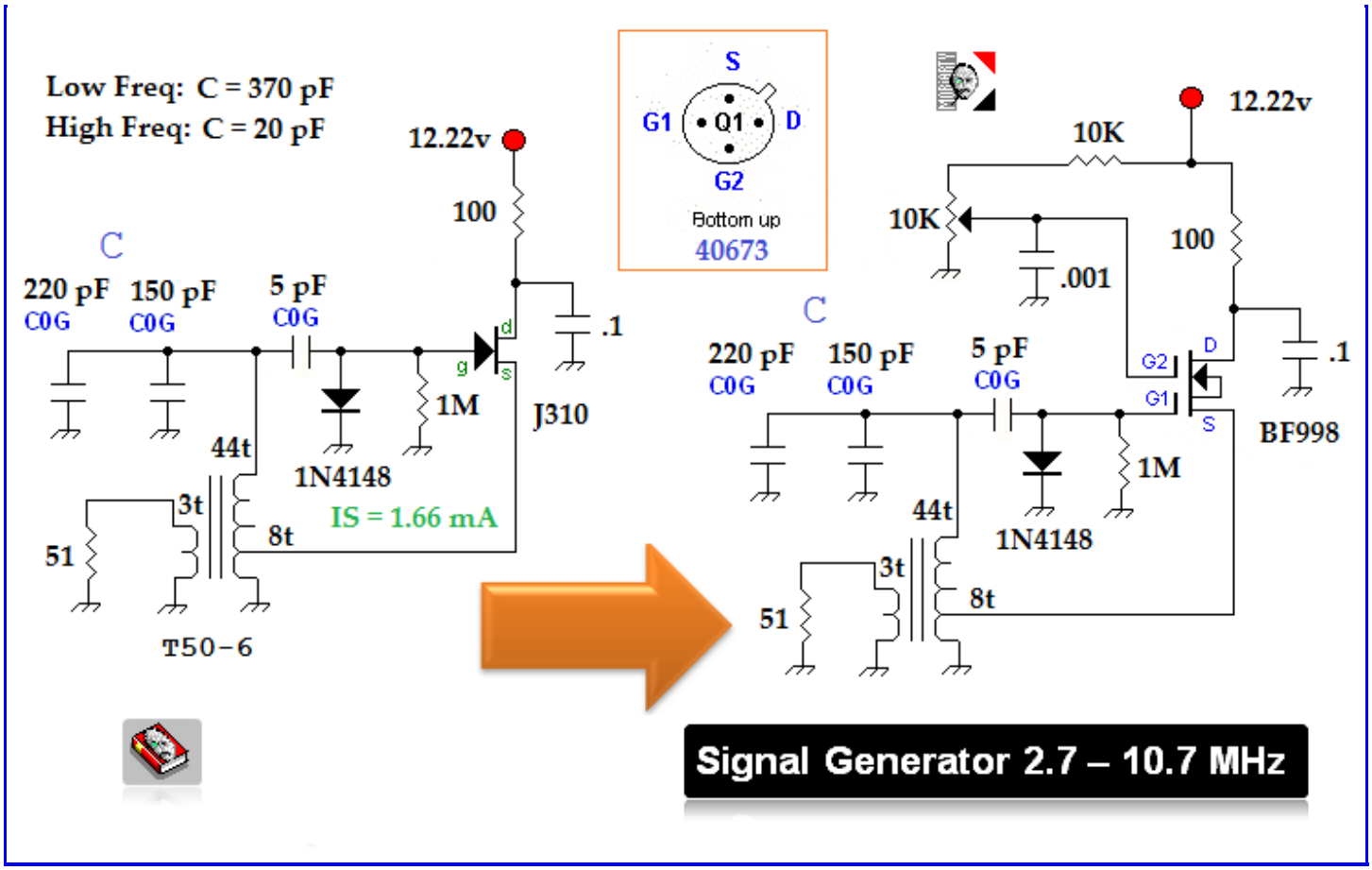
Above — I built a general purpose ~2.8 to 10.8 MHz signal generator (SG) for my lab in 2011. The first VFO topology tested was the Vackar. In my version, while employing a 100K ohm resistor as the buffer, the VFO only tuned from about 4 to 8 MHz and suffered from *extreme* amplitude variation as I changed the frequency across its range. For sweeping filters or measuring Q, a signal leveling circuit would be needed as normally we like our SG output to be flat across its frequency range. I later changed to a Hartley VFO because of its flatter output and the wider available frequency range with any given resonator.

This initial Vackar VFO experiment wasn't a total waste as I learned a way to accurately sweep a **D**evice **U**nder **T**est with an unlevel amplitude SG. Measure the peak-peak voltages of the DUT with a signal generator and an oscilloscope in the same manner we measure insertion loss or gain in a 50 ohm system: Measure the peak-to-peak voltage with the DUT in line; disconnect the DUT, insert a barrel connector and then re-measure.

The dBm difference between the 2 becomes the dB value to plot for *that* frequency. To a sweep a filter, say for example, a band-pass filter, find the center frequency and then sweep below and above that CF while plotting the dB versus frequency.
This "in and thorough" measurement described takes time, but resolves any frequency versus amplitude issues and can be used to test signal generators. We tend to ignore things like cable loss versus frequency and scope or spectrum analyzer ripple.

Still, it will be easier to just use a Hartley VFO where our sweeps are assumed to be level due to the flatness of the amplitude versus frequency for **small excursions** such as 3 dB band-pass filter sweeps. Do not expect amplitude flatness over wide excursions however — this requires additional circuitry.

To reduce noise and boost fidelity, this SG runs modest current and was not designed for battery use.



Above — My Hartley VFO is morphed into a double-gate MOSFET VFO; this was a mistake and I make lots of them.

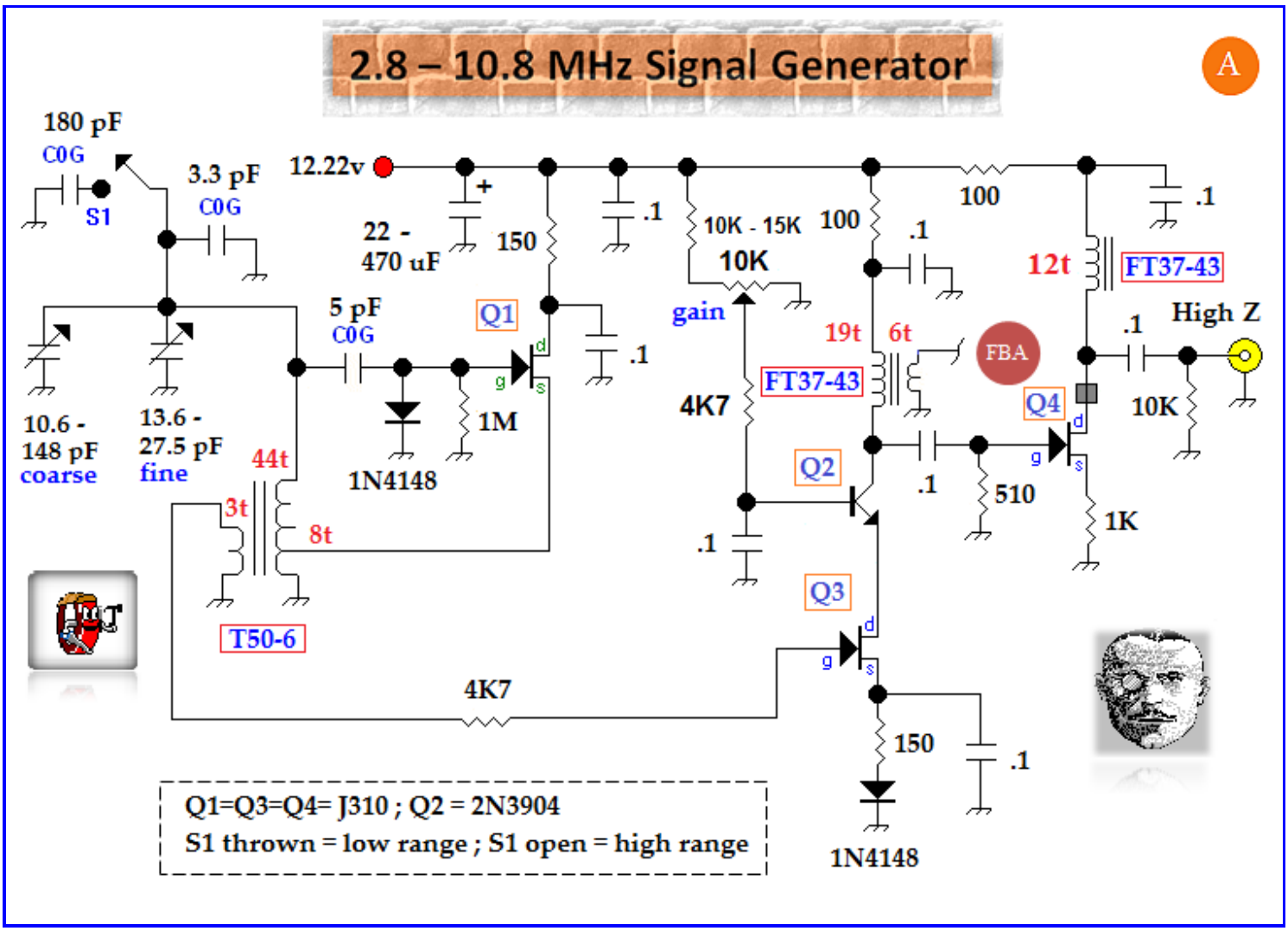
When venturing out, it's often best to confirm a proven design is working before morphing it to something untried. Shown above left is the Hartley oscillator from EMRFD Chapter 7 sans buffer. Fixed "tuning" capacitors; either 20 pF (not shown) or 370 pF (150 pF + 220 pF) represent the intended high and low frequency swing of my air variable tuning capacitors.

I wanted a variable amplitude VFO and thus replaced the JFET with a double-gate MOSFET using a simple variable voltage divider to control gate 2. I showed this to Wes, W7ZOI and he informed me that the flicker noise of MOSFETs precludes their use in oscillators. I have always wondered why I've never seen MOSFET VFOs in any radio literature.

Above — My project chassis fitted with hardware. I employed 2 air variable tuning capacitors — the fine tuning capacitor ranges 13.6 to 27.5 pF,

features built in 6:1 reduction gear and was purchased from Doug DeMaw many years ago. I secured the main copper board with 6 number 8 bolts. Rubber feet provide a stable, shock resistant base for the sheet metal box.

Final Build

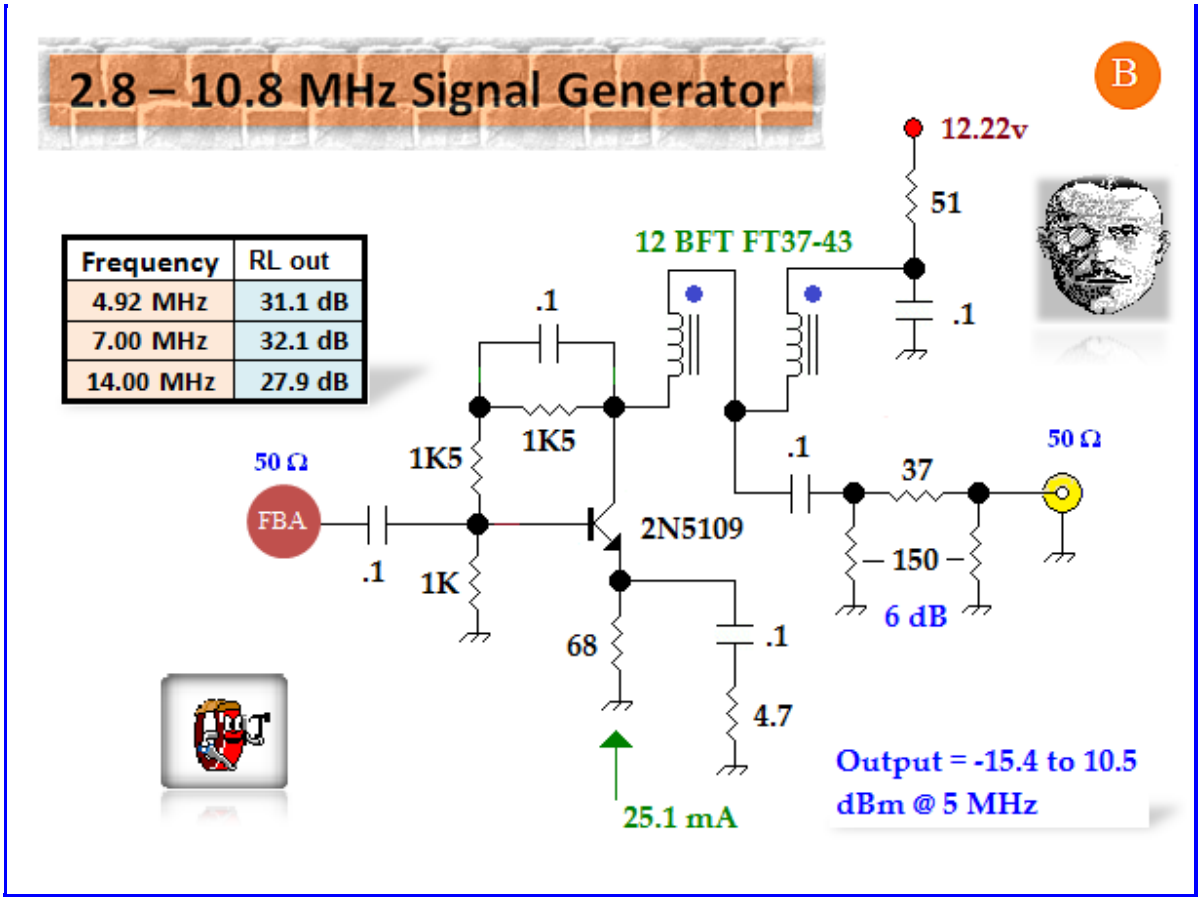


Above — Oscillator + buffer schematics of the latest version of my signal generator. I spent 1 evening playing with VFO designs and settled on the simple Hartley from EMRFD, Figure 7 .27. The 3 turn link provided lower distortion than coupling the oscillator to its buffer by the JFET source or gate.

Regulated 12.2 VDC powers the oscillator; avoiding the typical 5-9 VDC voltage regulator we normally use. A 22 to 470 µF cap should be employed to filter any voltage regulator noise from our DC supply. Mine has a 470 µF capacitor.

1 hour drift lies under 40 Hertz when averaged from 15 different frequency points between minimum and maximum. The Q2/Q3, Q4 and Q5 transformer inductances were optimized to allow good signal and/or matching performance in the ~ 2.5 to 10.8 MHz frequency range.

A hybrid cascode (hycas) buffer with variable base bias on Q2 forms the amplitude control for both the high impedance and low impedance outputs. The 510 ohm gate resistance on Q4 terminates the hycas amplifier and sets up a known output impedance to drive the 50 ohm feedback amp. I measured a greater than 22 dB return loss on the output of the 19:6 turn transformer from 4 to 14 MHz — indicating it drives the feedback amp reasonably well.



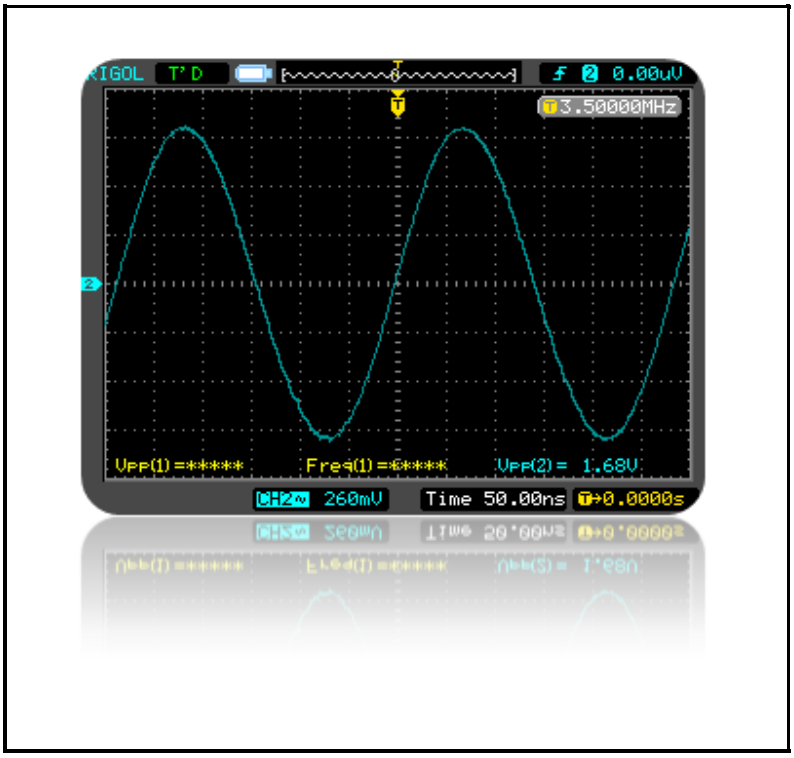
Above — The 50 ohm impedance feedback amp. Running 25.1 mA current allowed a clean sine wave output up to 2.12 volts-peak to peak into a 50 Ω terminated oscilloscope, plus an output return loss of over 30 dB across the SG tuning range. 3 tabled output return loss measurements are shown; including an out-of-range 14 MHz measurement.

Two series resistors made up the 37 ohm "resistor" depicted in the 6 dB pad, although a 39 Ω resistor would work fine.

At low output amplitudes, I typically stick an external 6, 10 or 20 dB attenuator on the output since the hycas amp can distort the signal a little when the gain control is set to a really low bias voltage on Q2. Then I fine tune the output power with the gain control.

Above — A "lid off" front panel photograph. Click or click for other photos. I'm now using miniature pots with a shaft diameter of 3.18mm. The potentiometer shaft lacks a knob and I'll purchase some on my next parts order.

Above — The completed signal generator.



Above — Signal generator output at 3.5 MHz.

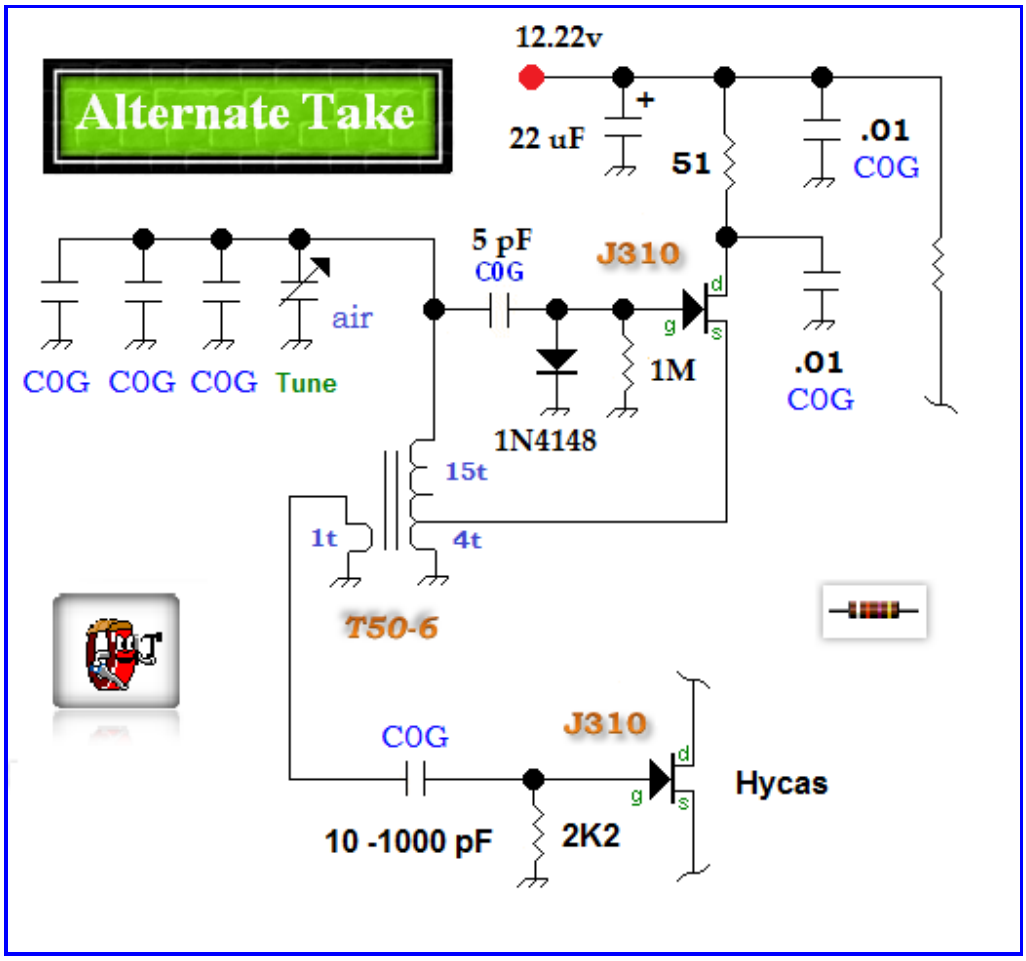
I appreciate that the VFO tank would be difficult to replicate since the 2 air variable capacitors are unique, however this is true of most VFOs. Wes wrote some great notes in EMRFD Chapter 7 regarding copying signal generators and the versatility of the Hartley VFO. I hope this project furnishes some ideas that spawn you to build something better than I did. *(I've received over 350 related emails since posting this page in 2011*

and many readers have built really great VFOs — Way To Go !)

QRP — PosData for December 17, 2013

I slightly boosted the tuning range from 2.8 to 10.8 MHz by dropping the 5 pF resonator capacitor to 3.3 pF in October 2013. The schematics now reflect this changes plus clarify 1-2 stumbling blocks readers had. For example: my use of a 100K gain pot plus a 150K maximum voltage limiting for the Q2 bias. My build still has these, however, I altered the schematic to show a common 10K gain pot, plus a fixed 10 to 15K resistor used to limit the Q2 bias to between 5 and 6 VDC maximum.

In reality, any reasonable pot and resistor will do since they function as simple voltage dividers. With a 12 volt supply, we don't want to drive the Q2 bias with more than ~ 5 to 6 VDC since this will just distort the AC signal as the hycas stage saturates. Measuring with a voltmeter, solder either a 10K, 12K, or 15K resistor to limit the maximum Q2 bias with the 10K potentiometer turned fully clockwise.



Above — Alternate way to couple the Hartley oscillator to the hycas buffer. Ground the JFET gate with a shunt resistor and lightly AC couple the JFET gate to the Hartley secondary coil with a series capacitor. You choose the capacitor value to limit the signal amplitude as needed. Our main goals are to lightly couple the Hartley tank to its buffer and avoid overdriving the hycas stage. Numerous examples of this "Alternate Take" circuit may be found on the QRP / SWL HomeBuilder web site.

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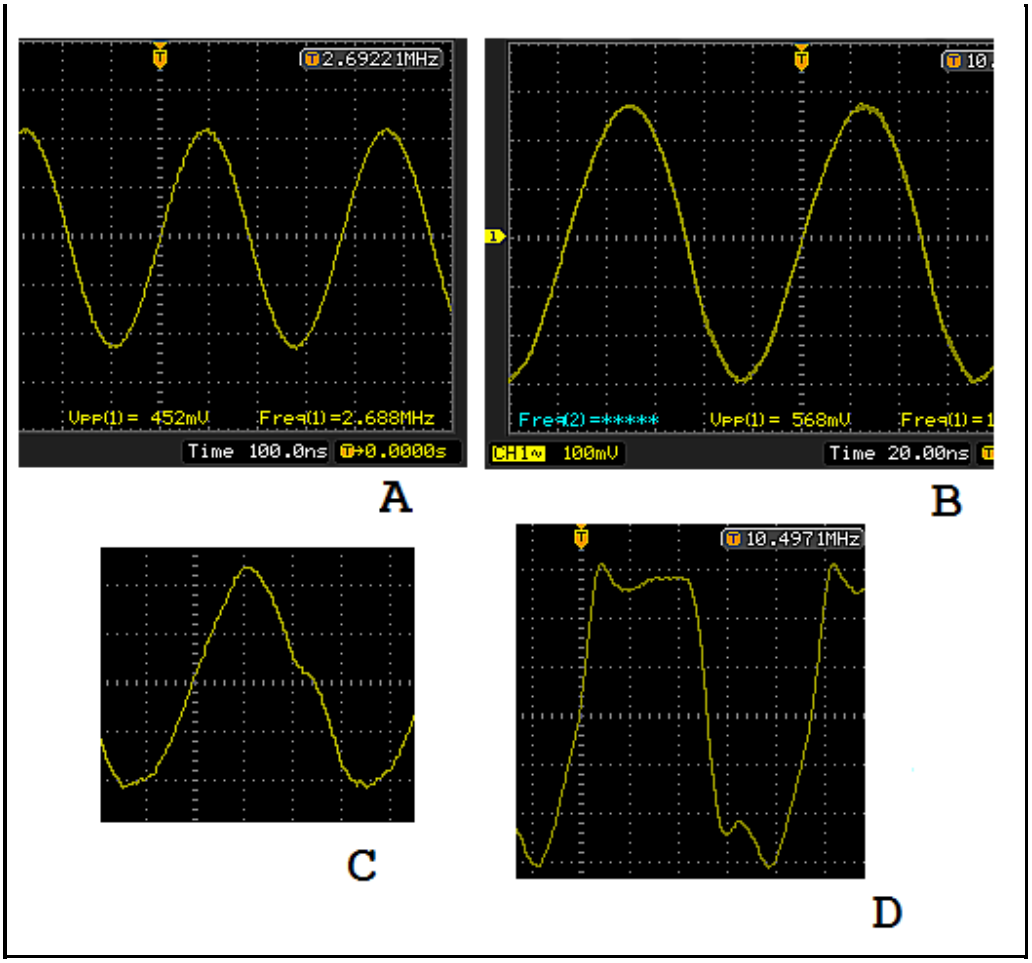
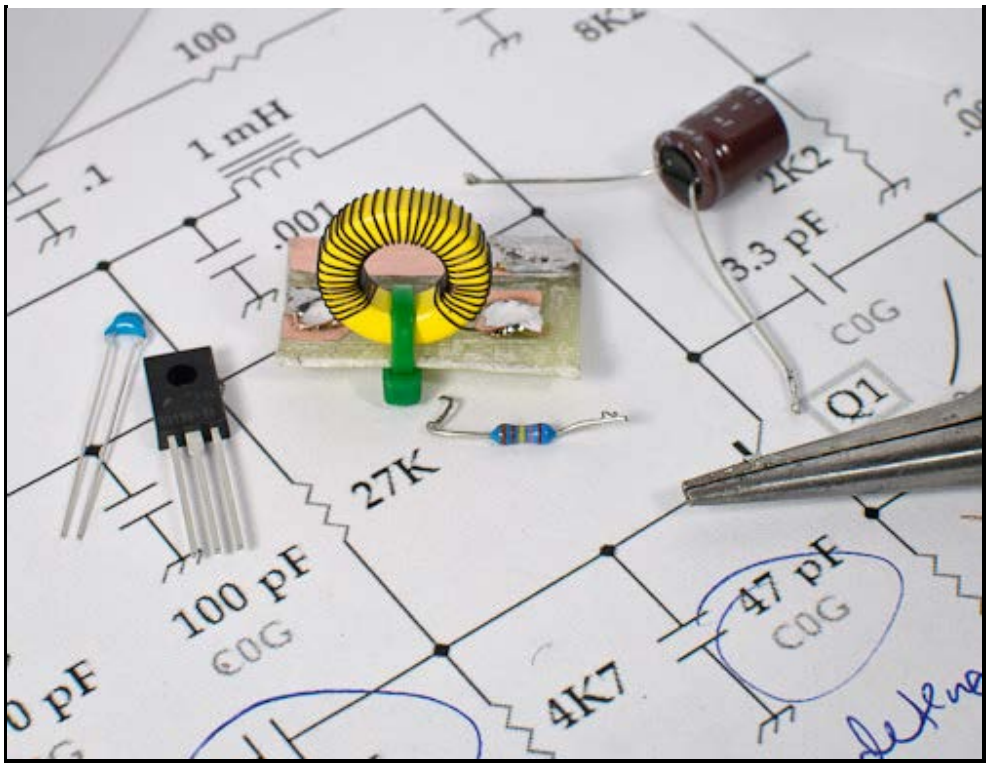
4. Miscellaneous Bits

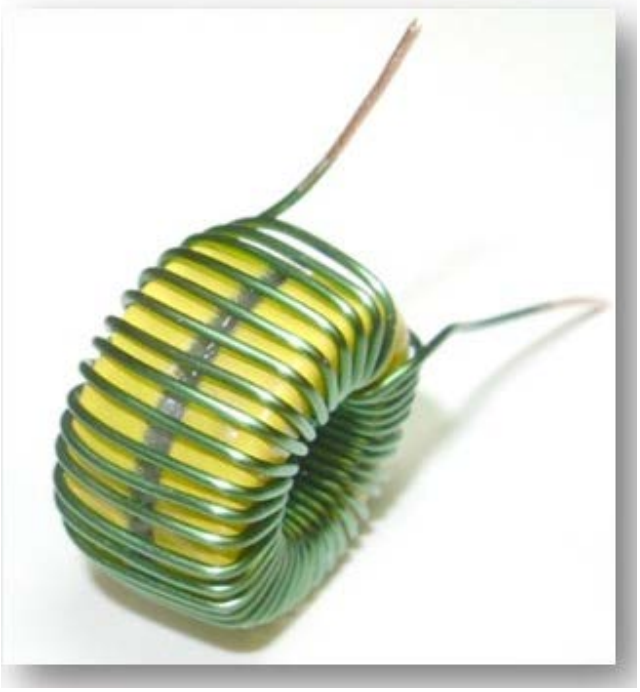


Above — Scope traces of an unbuffered Hartley oscillator with a X10 scope probe across a 51 ohm resistor across the 3 turn link. The unbuffered Hartley sine wave isn't harmonic free, but cleans up when properly buffered with a higher impedance amplifier. Figure A = the lowest frequency (2.7 MHz) — the distortion increased with frequency (Figure B was measured at 10.5 MHz). Figure C illustrates how slightly stronger output coupling with a 6 turn link trashes the output waveform — the strategy of using 2-3 links over the center of the main inductor works well.

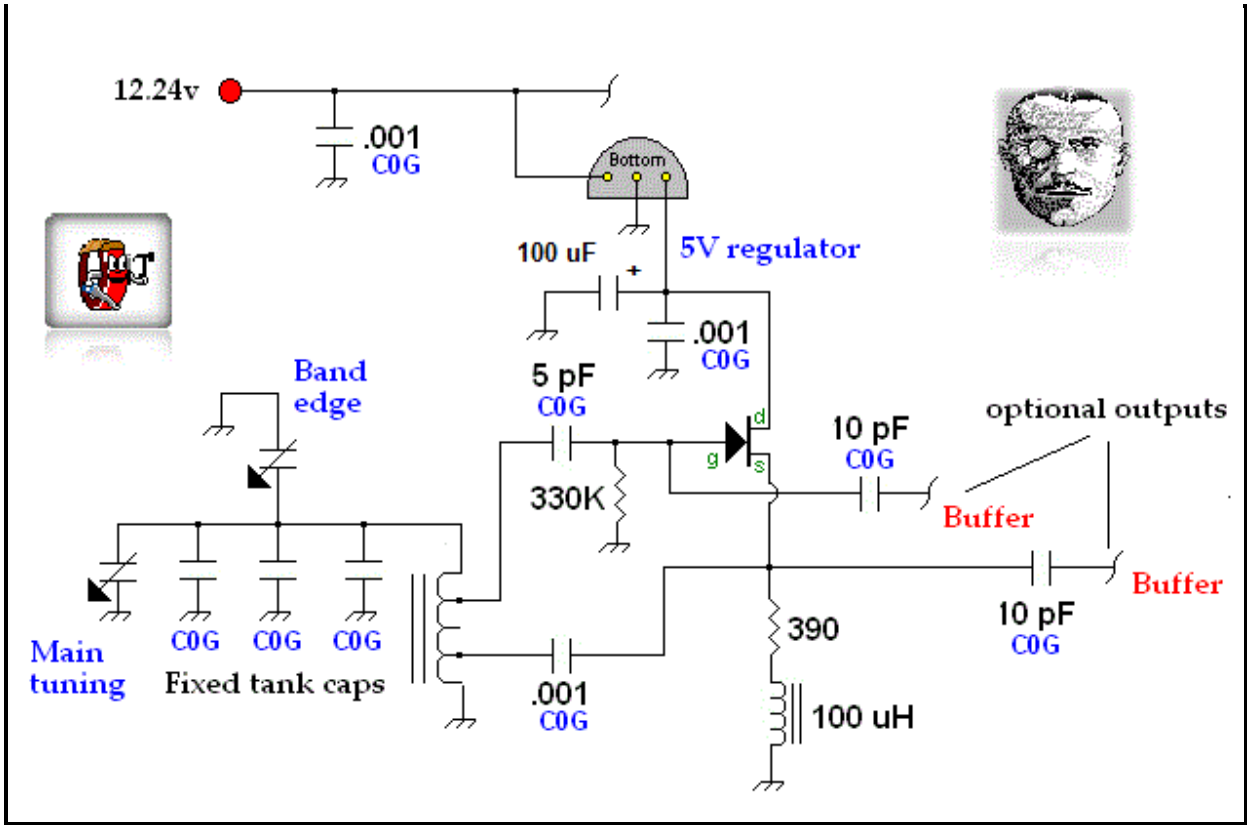
Figure D is the 6 link coupled Figure C oscillator with the gate clamping diode removed; yikes! I spent 4 hours studying what different current, voltages, coupling and so forth do to the Harley oscillator. I recommend the Hartley topology because it's simple, always starts and versatile.



Above — In my various signal generator experiments, I zap strapped the toroid to a small piece of thick copper board that was later soldered to the main board. The L seems robustly secured.

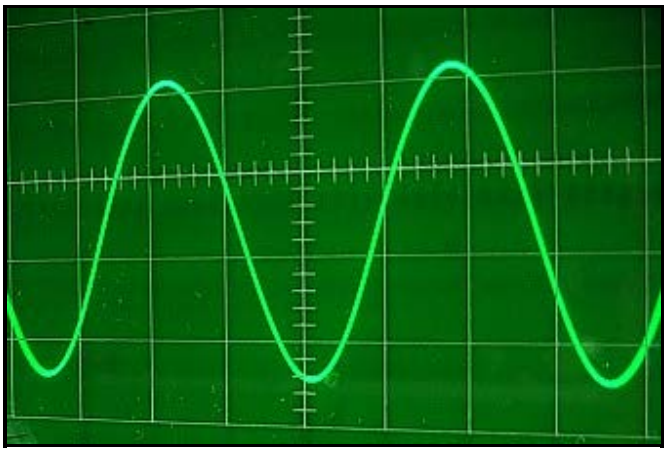


Above — A stacked toroid from the deleted VFO-2008 web page. I incorporated some of the information from the VFO-2008 page into this web page

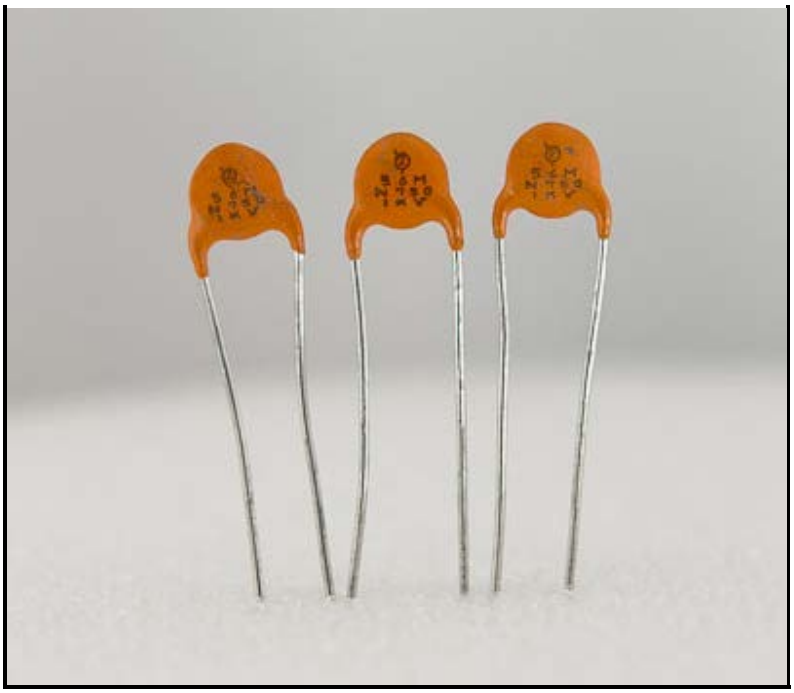


Above — In order of preference, 3 ways to couple a Hartley oscillator to its buffer. From now on, I'll couple with a 1-3 turn link since it gave a lower distortion signal than with source or gate coupling. This figure omits the gate clamping diode seen earlier— tapping the inductor as shown keepings the FET gate AC voltage at a reasonable level when not using a gate clamping diode. Some builders leave off the gate clamp diode that clips positive signal peaks for lower phase noise. The diode acts as an AGC and offers benefit. Reverse biasing this diode was suggested by Dr. Ulrich Rohde: see — Key Components of Modern Receiver Design - Part 2: Dr. Ulrich Rohde, KA2WEU , QST for June 1994.

A formula to use for the inductor taps: Divide the total turns by 1.45 to get the first tap and by 7.25 to get the second tap (near ground link).

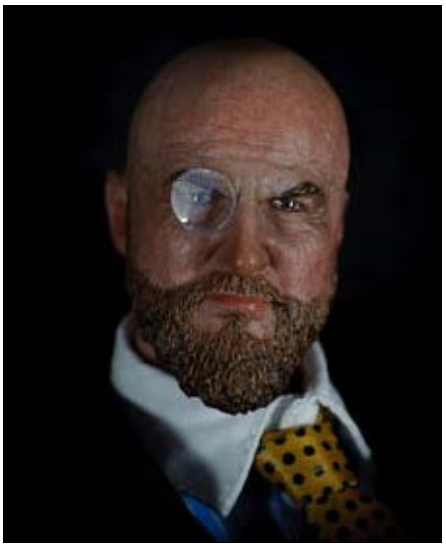


Enjoy your VFO experiments.



ON
HI-Z
50 OHM
GAIN
HI

The Butler Did It ! - First VHF Experiments 2011



Venturing into VHF, I felt like a beginner with no experience or confidence — however, the excitement and allure of new parts and circuits kept me going.

After performing a literature review, talking to some colleagues and renovating my QRP work bench, my first tasks involved buying some VHF parts + exploring the Butler crystal oscillator. Specifically, I'll cover my experiences with the common base version of the Butler oscillator.

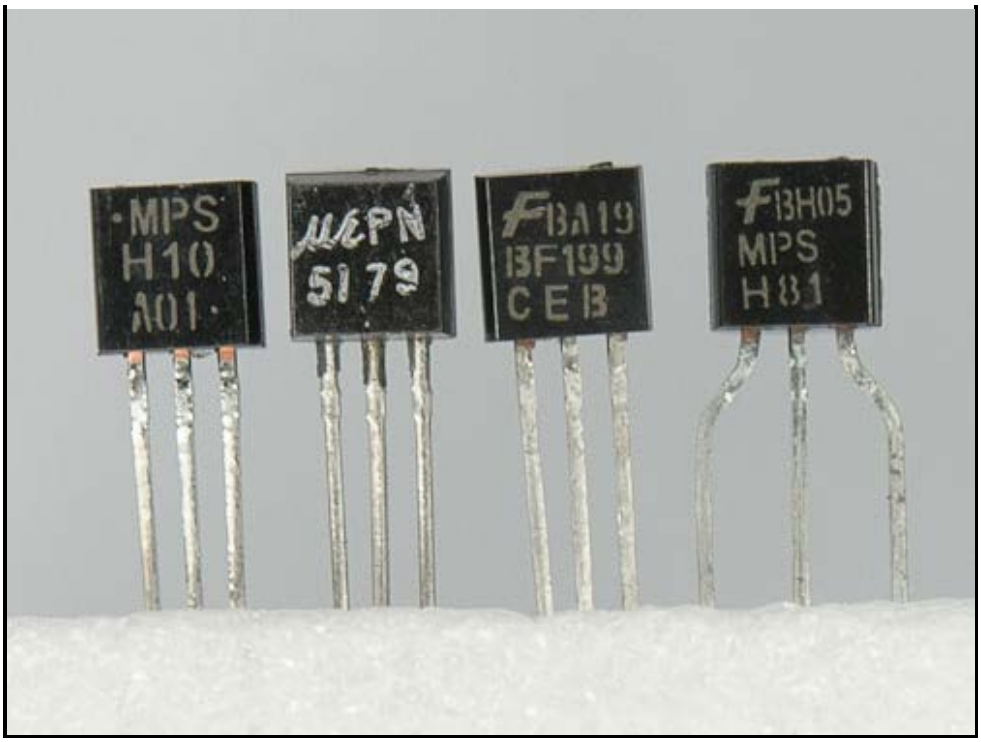
Like HF, the VHF knowledge base contains ever-present lore. Consider the Butler oscillator — I have read arguments stating that the emitter follower version of the Butler oscillattor is vastly superior to the common base version because the latter is prone to UHF and other spurs. These comments seem to have originated from a good book entitled Crystal Oscillator Circuits. *Revised Edition* by Robert J. Matthys published in 1992 by the Krieger Publishing Company.

While examining the schematics of professional/world class gear using a Butler, the common base version clearly dominates. Spectrum analysis and other measurements indicate that when common UHF oscillation management techniques are applied, common base Butler crystal oscillators work well.

Suppressing UHF oscillations with ferrite beads (and small value resistors), feedback, neutralization, limiting gain, etcetera are routine practices for us experimenters applying active devices that have strong gain into UHF on up. This is vanilla, or matter-of-fact construction for us; no worries. While fun and often convenient, lore ultimately stifles our progress.

Increasingly, I'm adopting the philosophy of Bob, K3NHI; "TMITK" — to measure is to know. Consider, too, you have to know what to measure and possess the required gear.

For BJTS, I purchased PN5179, MPSH10, 2SC3355, BF199, 2SC3357-RE, 2N5109/2N3866 and NE46134. The PNP part = MPSH81.



For JFETs, the J310 in TO-92 and SMT will remain my workhorse FET part along with a couple of other *low noise* JFETs and 2-gate MOSFETs. At VHF, the noise figure in a receiver chain is established by the first amplifier so a low noise preamp ranks important.

A collection of 100 volt NP0 capacitors ranging from 1 pF to 22 pF were added along with some chip and SMT caps as low as 0.5 pF. You might need a few air trimmer capacitors with a minimum capacitance ~ 2 pf — I applied 2 - 20 pF trimmer capacitors in most of the circuits that follow.

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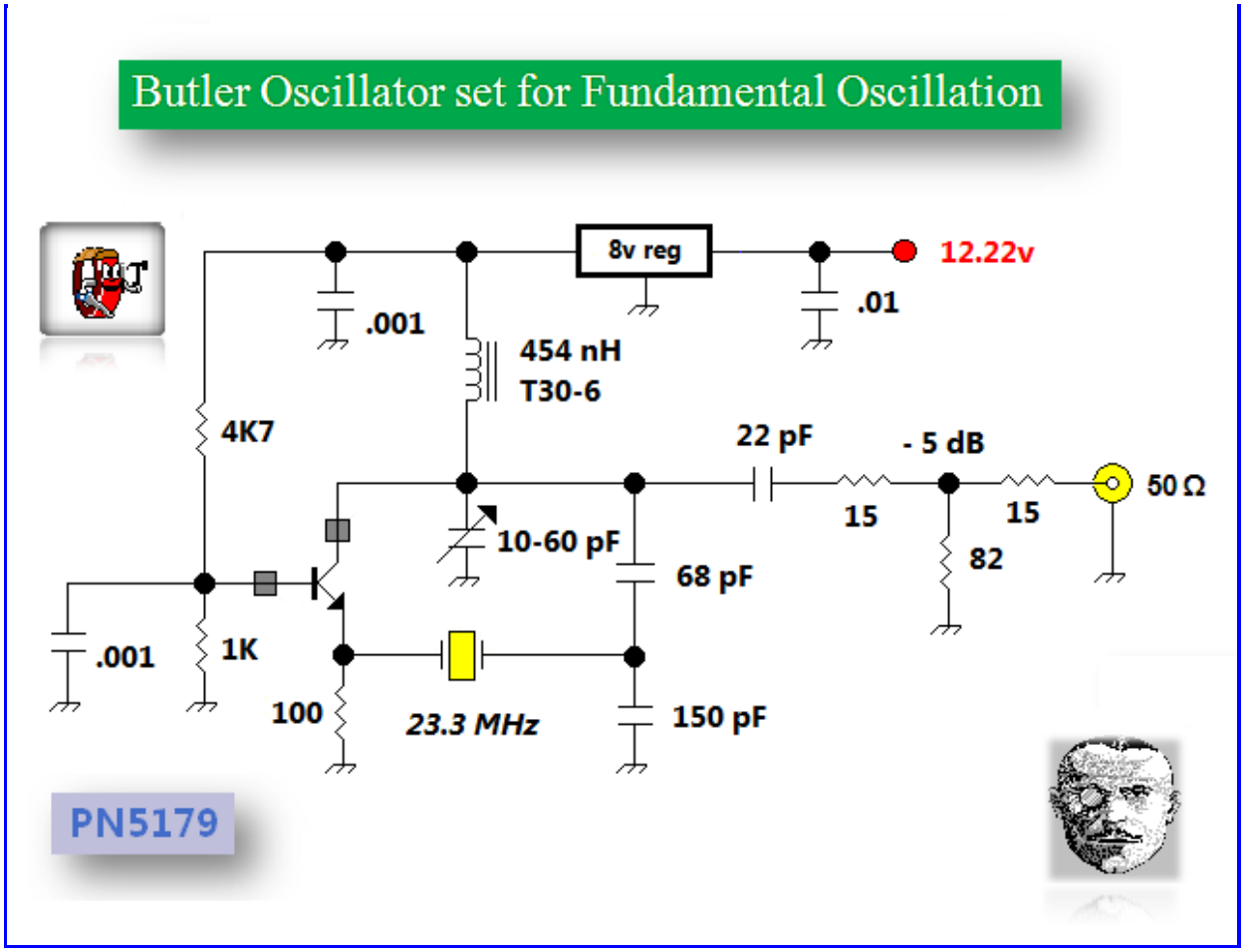
1. Experiments with a Butler Oscillator with a 23.3 MHz Fundamental Crystal

Fundamental Frequency

Long ago, I pulled a crystal marked 70.00000 MHz from a Drake Transmitter. I keep a fundamental oscillator based upon EMRFD Figure 4.23 on hand and verifed the fundamental frrequency at 23.3 MHz.

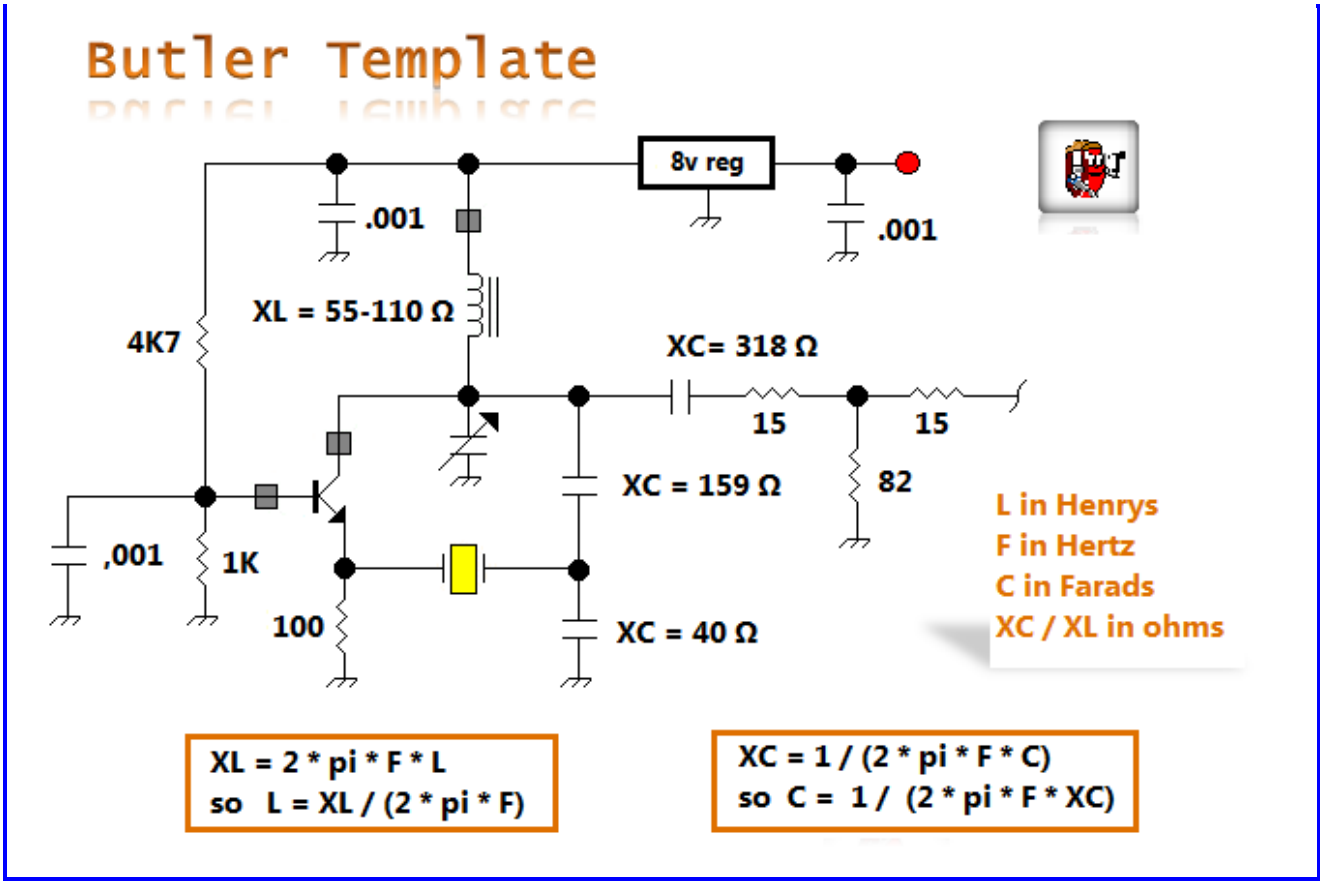


Above — A Butler oscillator arranged for output at the fundamental crystal frequency. While commonly arranged as an overtone oscillator, the Butler is a good oscillator for any application. Consider, for example, EMRFD Figure 7.32. Wes applied the Butler at a 14 MHz fundamental because he wanted the lowest phase noise and IMD prone signal source possible. Tellingly, his buffer circuitry also conforms to this high standard.

My initial waveform looked distorted and prompted a solution. My experience yields that the L value needs to be adjusted for the best looking waveform in the Butler circuit. The inductor wire, wrapped around a T30-6 toroid was either scrunched to increase the inductance, or expanded to decrease the L while re-peaking the trimmer cap. Eventually, with patience, a beautiful sine wave emerged on my 'scope. I removed and measured the L with an ADE inductance meter. Consider all of my reported inductance values as nominal — gentle expansion or contraction of the inductor coils might be required to get an agreeable sine wave.

Click for the oscillocope tracing at 23.3 MHz.

The Butler at Overtone Frequencies



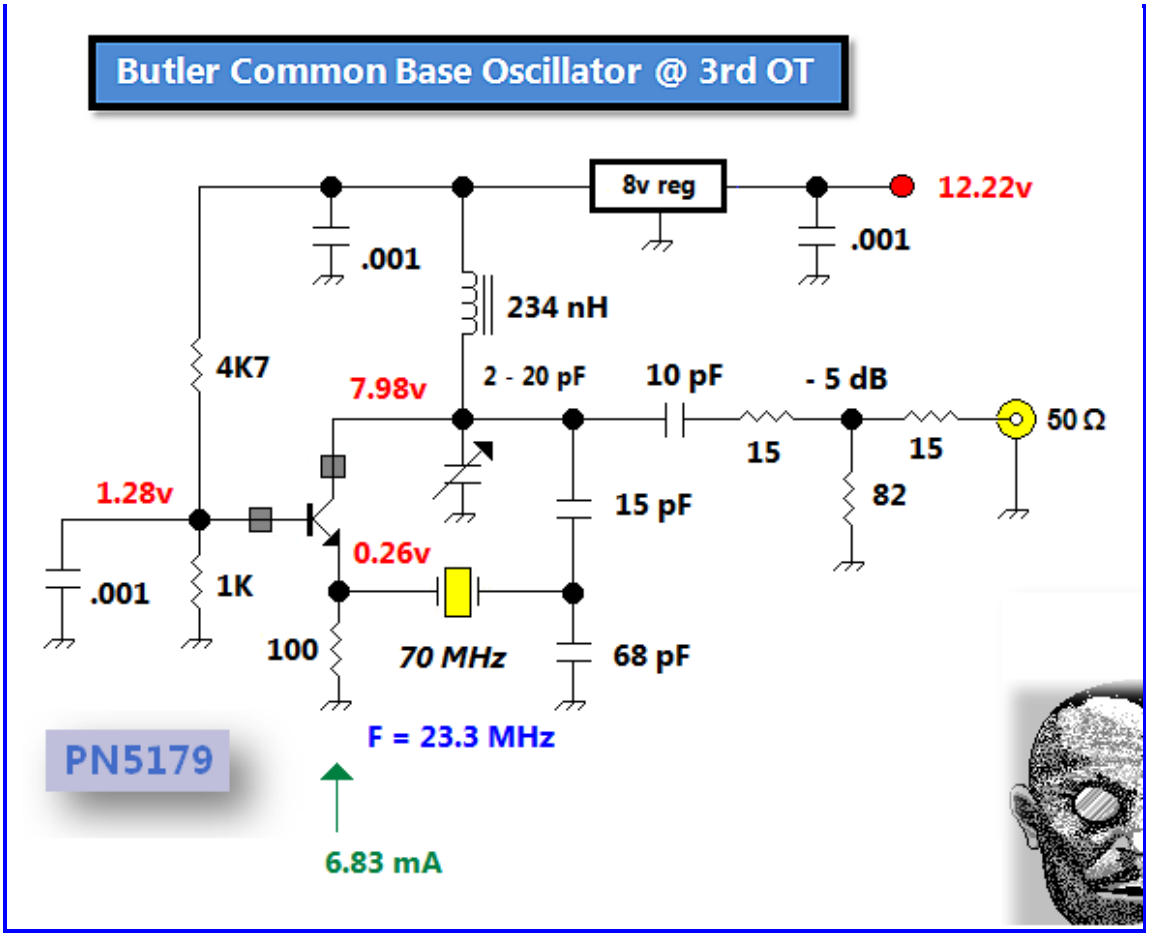
Above — A template (of sorts) for calculating Butler capacitor and coil values. The concept, rather than the absolute value matters most. I examined some well-designed Butler oscillators from professional equipment and determined their average XL and XC values. From the reactances shown, calculate the L and C values for the overtone frequency of interest with the 2 formulas in orange boxes. *Remember these XL or XC values just serve as starting values for experiments.*

For example : At 50 MHz with an XL of 108 Ω : L = 108 Ω / (6.28 * 50000000 Hz) = 0.000000344 H or 344 nH.

Fine tuning of the capacitor and inductor values might be required since factors including buffer input resistance +/- reactances, the overtone frequency and/or your breadboard layout may affect your Butler oscilator function.

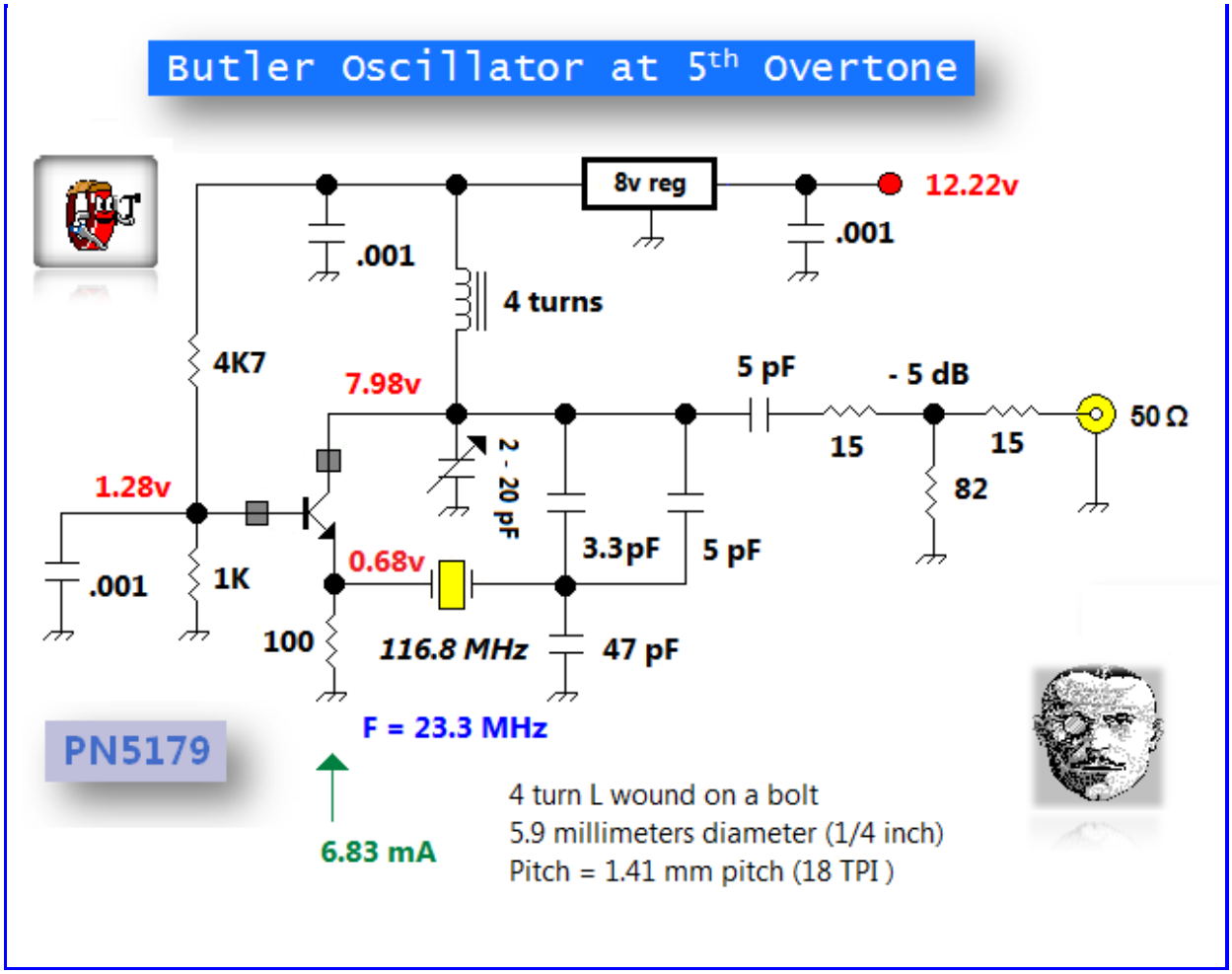
In the experiments from the 3rd to 9th overtone frequency, the L = an air inductor wound with 21 to 22 AWG wire on a bolt.

Third Overtone Frequency



Above — The Butler oscillator now arranged for output at the third overtone. Click and click for the 'scope outputs at the third crystal overtone. The first scope tracing was slightly mistuned. Once again, the inductor had to be gently squished or contracted to obtain a pristine sine wave.

Fifth Overtone Frequency



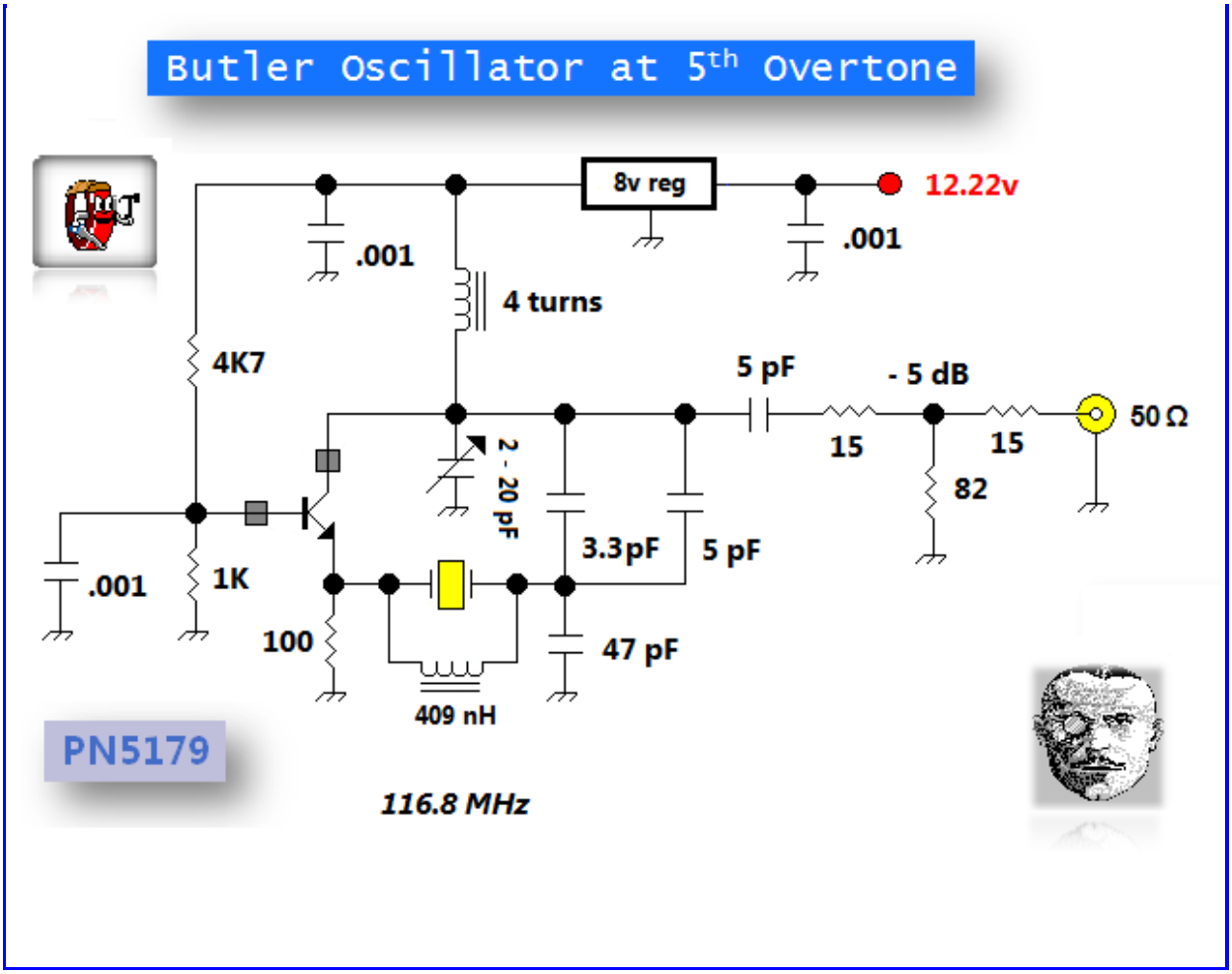
Above — The Butler oscillator at the 23.3 Mhz crystal's 5th overtone. Here's the 'scope tracing and an Ugly Construction photograph. The output = 7 dBm — very cool.

Some authors directly connect the attenuated output to the Local Oscillator port of a diode ring mixer. The signal is adjusted to the desired 7 dBm power by tweaking the 50 Ω pad attenuation, and/or the NPN's current. A good example = Single-Conversion Microwave SSB/CW Transceivers by Rick Campbell in QST for May, 1993.

The circuit above was measured with a 50 Ω terminated oscilloscope however, I also tested it with a 10X probe attached to termination resistors from 51 to 1 Meg ohms. When changing the 51 ohm termination resistor to a higher value such as as 47K, a previously working Butler may stop oscillating. The buffer input impedance and capacitance greatly affected the oscillator in my experiments.

When I wired up a Butler, 1 of 3 things happened: it did not oscillate, it gave a distorted output waveform, or it wowed me with a nice sine wave. Tuning the L-C tank is critical + finicky and may test your patience.

Adding an Inductor Across the Crystal



Above — The 5th overtone Butler with an inductor in parallel with the crystal

Above — My 116.8 MHz oscillator breadboard with an inductor wound on a T50-6 with wide spacing to allow scrunching and expanding of the windings across the crystal. I roughly determined my L should be ~ 400 nH and wound this on a # 6 toroid. While observing the output in my scope, I scrunched and then expanded its windings and adjusted the trimmer capacitor. The goal was to find a clean signal that snuffed out immediately when the trimmer cap was tuned off resonance. After finding the optimal L, I later removed and measured the coil. There is very little "wiggle room" — the oscillator tunes up and then dies very sharply as you tweak the trimmer cap. No sidebands were observed.

The inductor across the crystal is optional — some suggest it might only be needed above ~ 70 MHz.

At frequencies above ~70 MHz, the parallel capacitance of the xtal (C par) approaches the internal series resistance of the xtal and this provides an alternate path around the crystal for the signal and may short-circuit the crystal. The parallel inductor resonates with the crystal's parallel capacitance and tunes it out, so the crystal remains unbypassed.

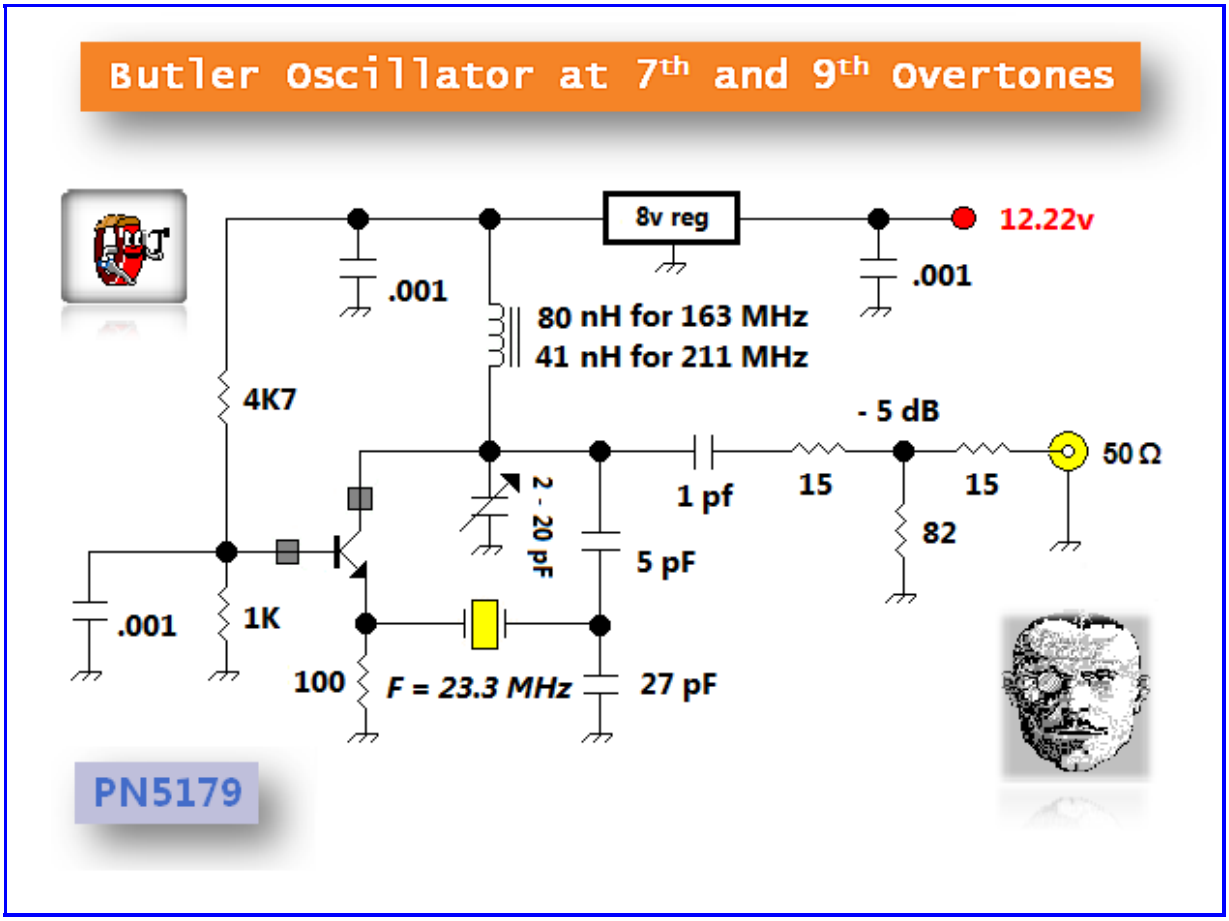
Our teacher, Wes, W7ZOI published a document covering the Butler parallel crystal inductor here.

Simply put — the inductor allows clean tuning and output. That is, when you tune the trimmer to one side or another, the oscillator just dies and doesn't produce the sidebands that are shown in Wes' web article.

Since many of us choose computer, or other surplus crystals, a high C par + low Q crystal might give you tuning woes depending on your overtone frequency. In this case, adding the L to your circuit may improve tuning and ward off any unwanted sidebands.

With my particular crystal the parallel inductor is not needed, however, I can report that even slight mistuning just snuffed out my oscillator with the added inductor.

Seventh and Ninth Overtone Frequency



Above — I decided to take the Butler up to the 7th and 9th overtones. The 1 pF coupling capacitor proved the most critical part; for example, if I raised it to 5 pF, the oscillator would not tune above the 5th overtone. I wound the inductors on bolts and then compressed or stretched the links to get the perfect inductance. At these frequencies, stray inductance becomes quite significant and my coils were 1 to 1.5 turns less than indicated on

a spreadsheet coil inductance calculator. Click for the seventh overtone 'scope shot. Click for the ninth OT 'scope tracing.

I'm confident that if my 'scope bandwidth was higher, I could have resonated the 11th overtone.

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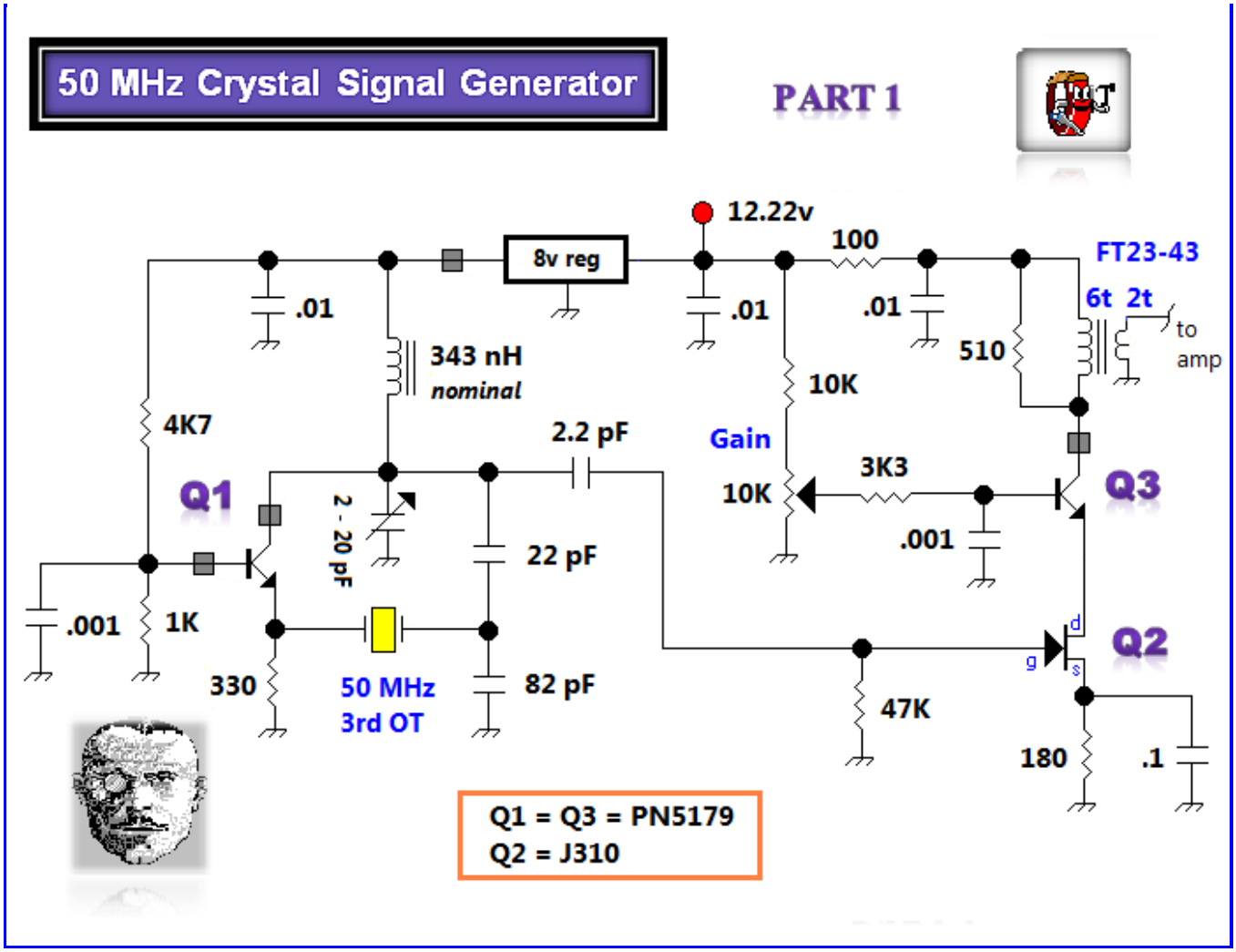
2. Butler Oscillator-based 50 MHz Signal Generator

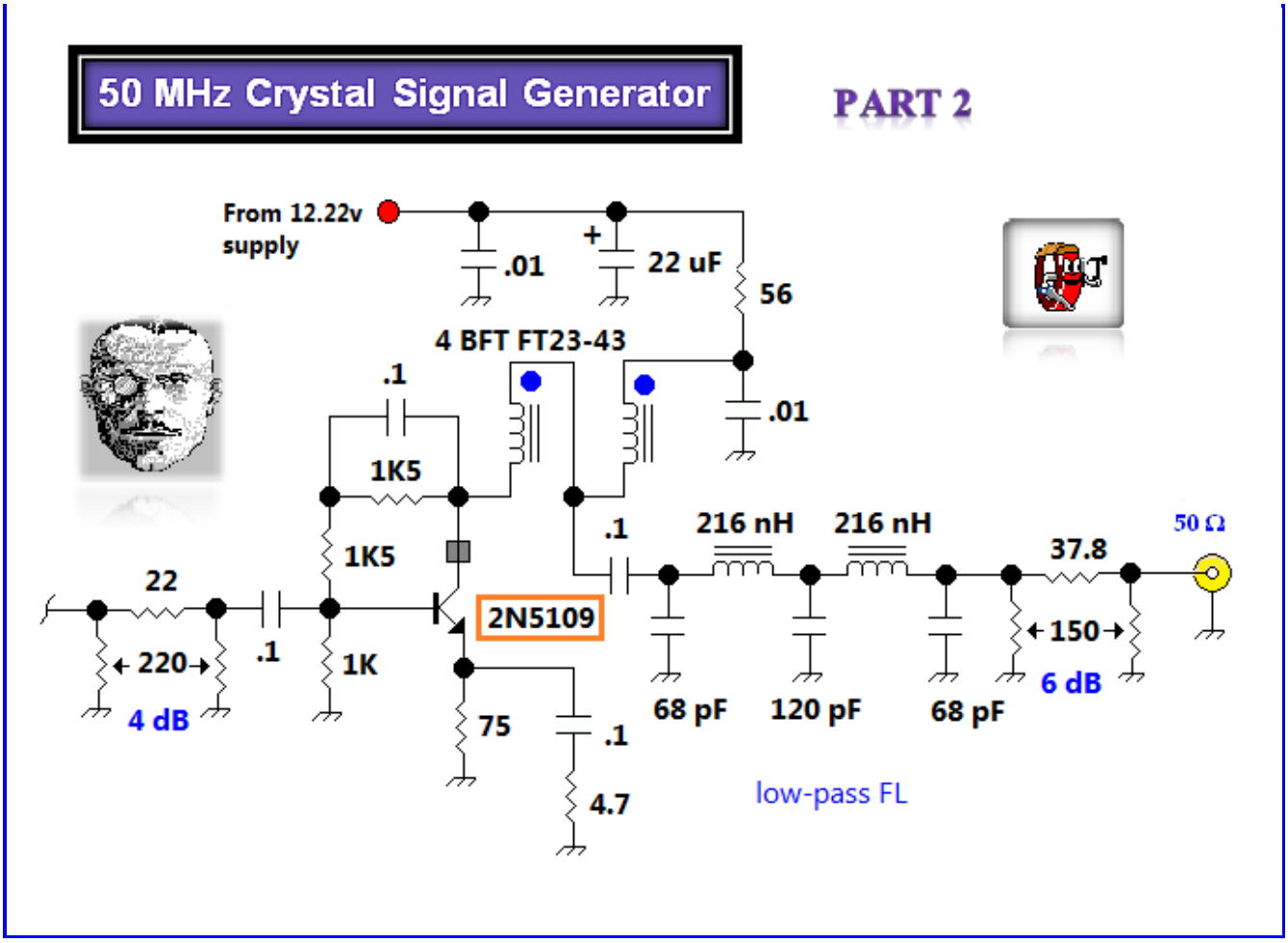
Lacking a 6 Meter band signal generator, I decided to build a 1 frequency device employing a Butler oscillator. Some may laugh at a 1 frequency signal generator — I won't since I'll use it to design and align amplifers, filters, a new 6 Meter band VCO and measure scattering parameters. Besides, you can double, triple, VXO or mix single frequency generators with another variable oscillator — this web site has roots in humble, simple test equipment.

In my bag, I found a crystal labelled 50.0000 MHz and measured its fundamental at 16.67 MHz in a simple Colpitt's oscillator — perfect . Design goals included variable amplitude, strong reverse isolation + reasonable shielding and return loss. Since, I lack another 6 Meter band signal generator I could not measure return loss, however, choosing proven circuits + a terminal attenuator pad will help.



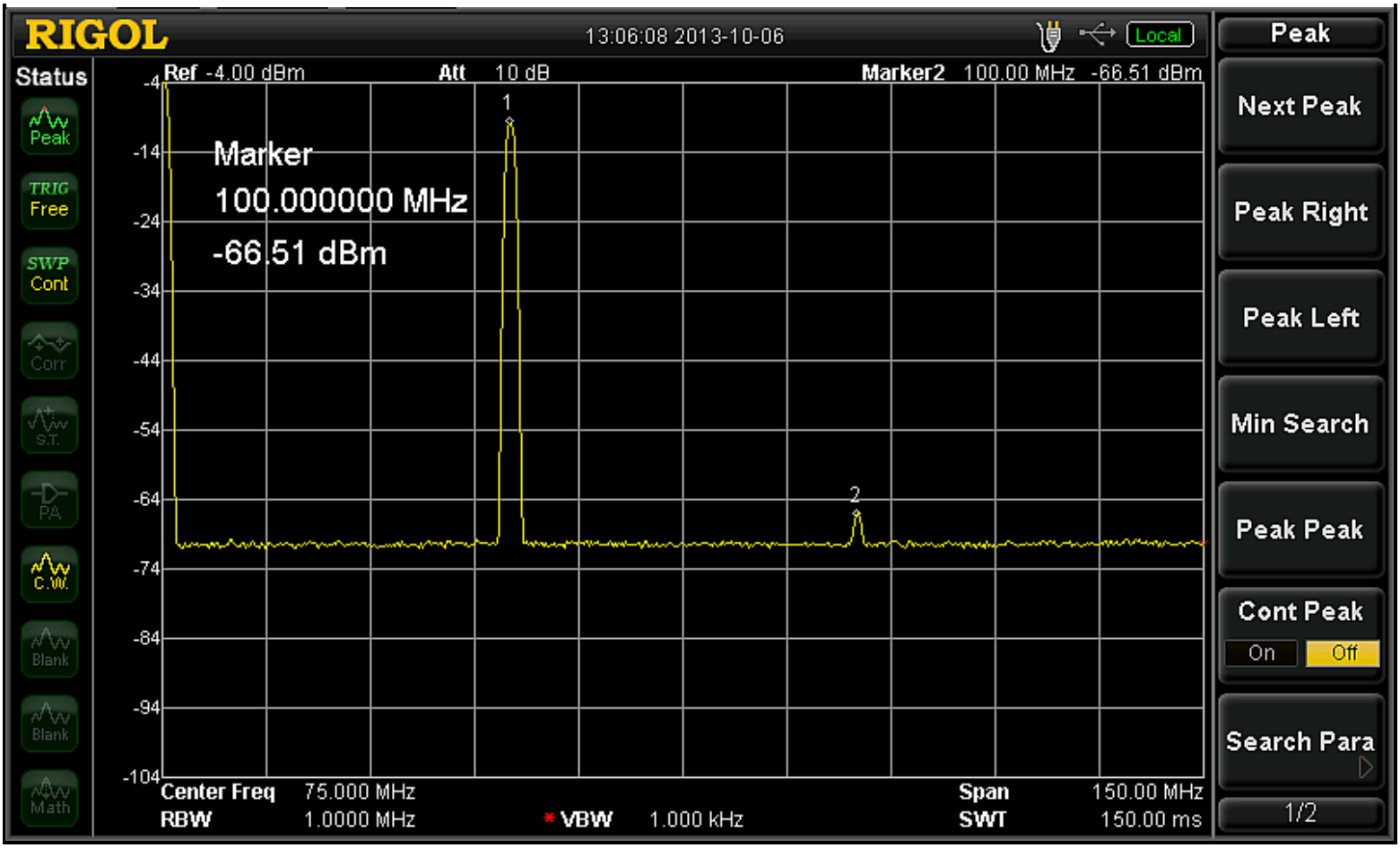
Above — Butler oscillator + hycas buffer. The 22 pF cap in the Butler was originally 15 pF, but when tested with various coils, oscillations proved a litle sluggish, so the 22 pF was substituted. I experimented with the coil, but ran out of time, so I wound a few turns of wire on a T50-10 toroid and soldered it in. After some careful manipulation of the windings, a glorious sine wave arose. I removed the L, measured it at 343 nH and then re-soldered it in place. An air coil or other size 6 or 10 powdered iron inductor should work fine.

Click for a moderate resolution photograph of the entire project.



Above — The final amplifier and low-pass filter. The maximum output of the Q2/Q3 amp is hot and can overdrive the 2N5109, therefore a 4 dB pad was added. This pad also improves the input return loss of the 2N5109. An alternate technique might be to further reduce the maximal DC bias voltage into the Q3 base and remove the 4 dB pad.

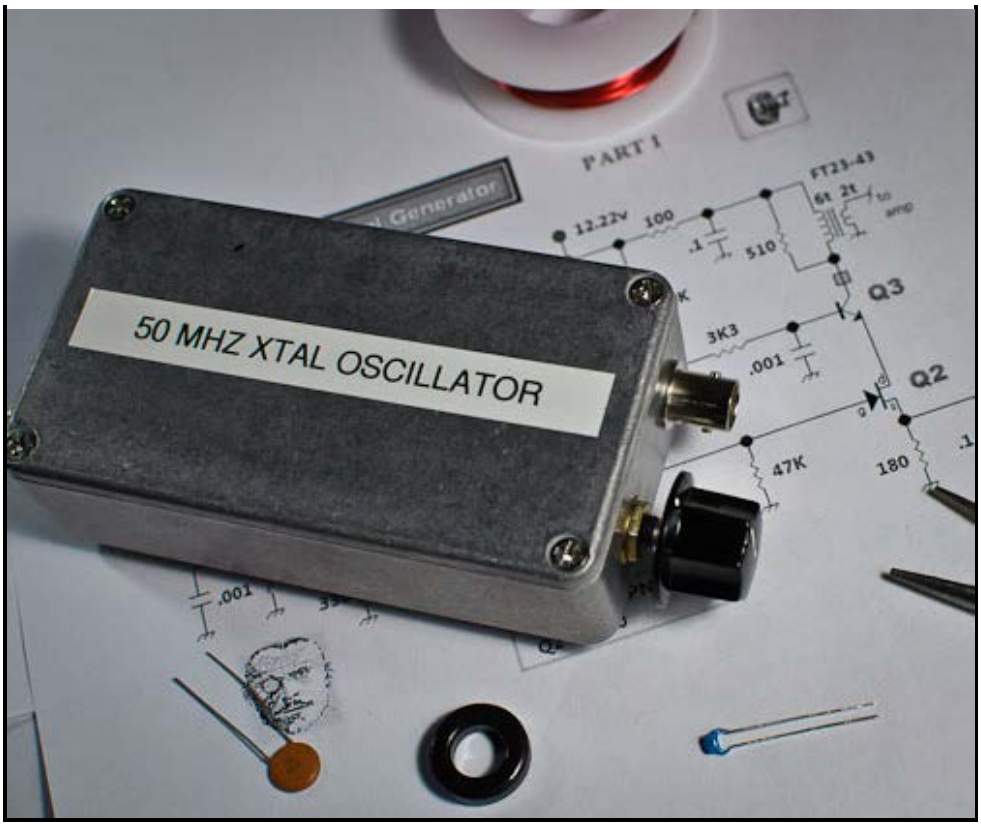
Click for the minimum and maximum output voltage waveforms from my 'scope



Above — Spectrum analysis with the generator output at -10 dBm. I placed Marker # 2 on the 2nd harmonic peak (100 MHz) and measured the power. The low-pass filter only attenuates the 2nd harmonic by ~ 25 dB, however, with the nice sine wave from the Butler, the 2nd harmonic measured -66.51 dBm = -56.51 dBc.

The 37.8 resistance indicated in the pad = 2 resistors in series. Many just use a 37 or 39 Ω R.

I look forward to advancing my VHF skills with this little signal generator.



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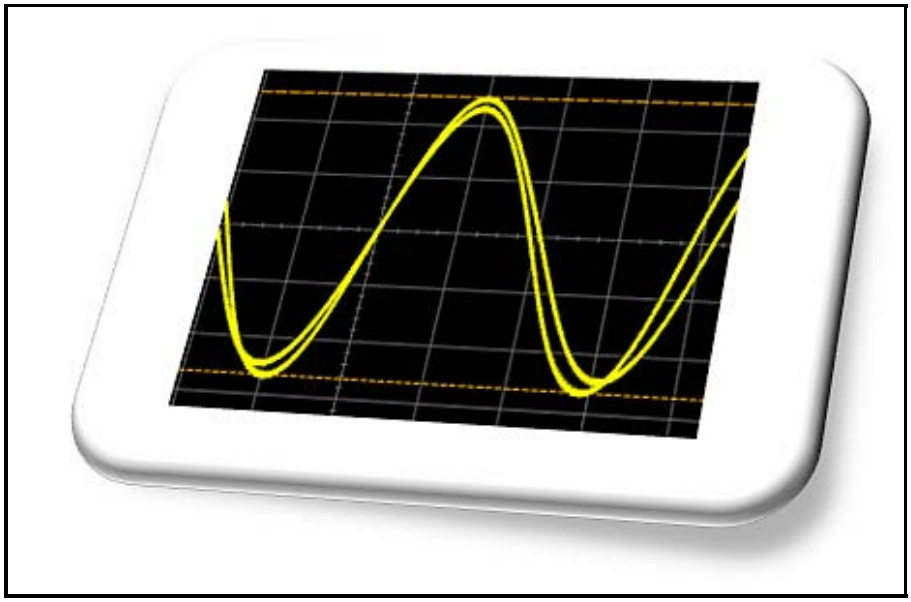
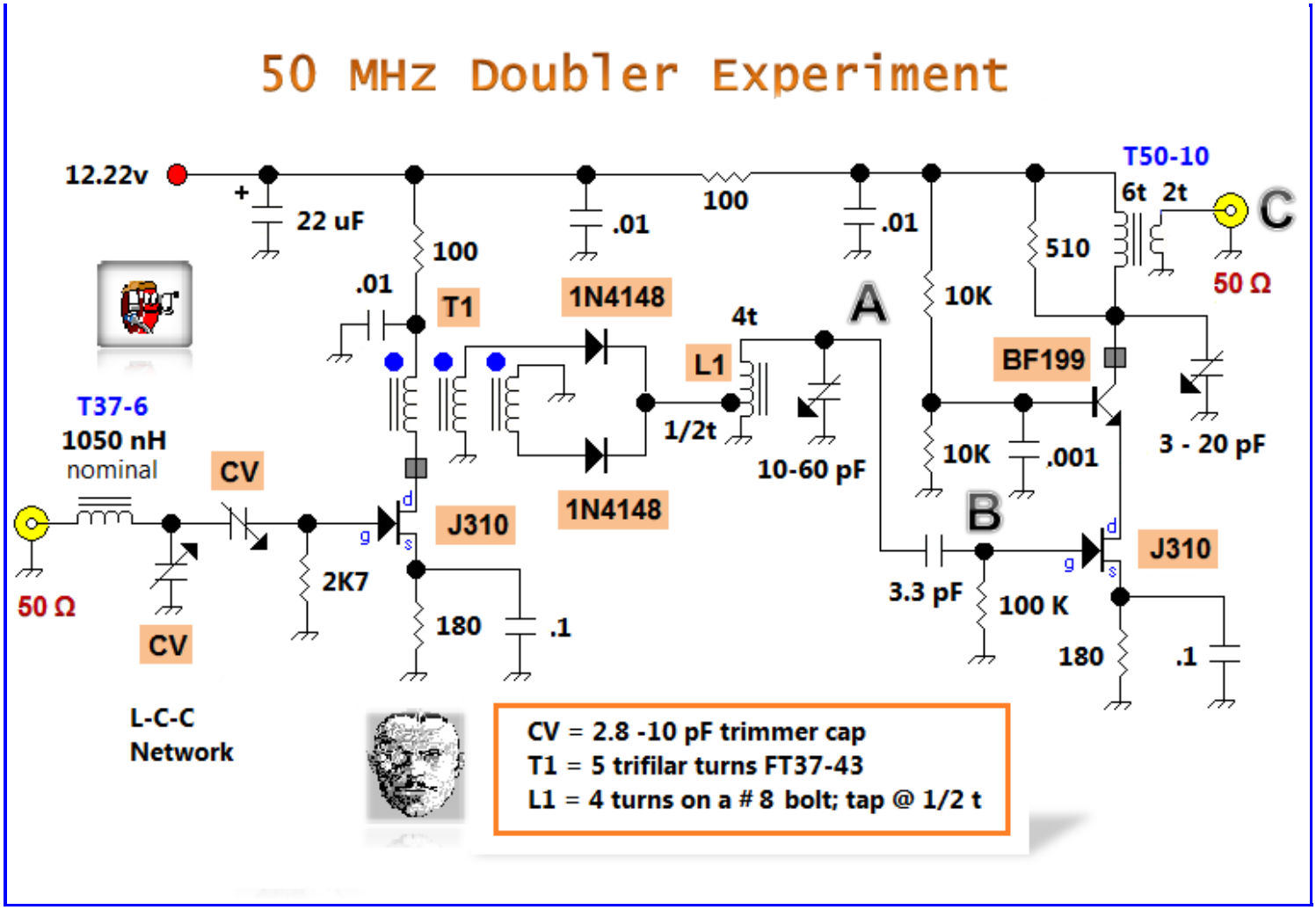
3. Doubler for the 50 MHz Signal Generator





Above — An experimental frequency doubler for the Butler Oscillator-based 50 MHz Signal Generator shown above.

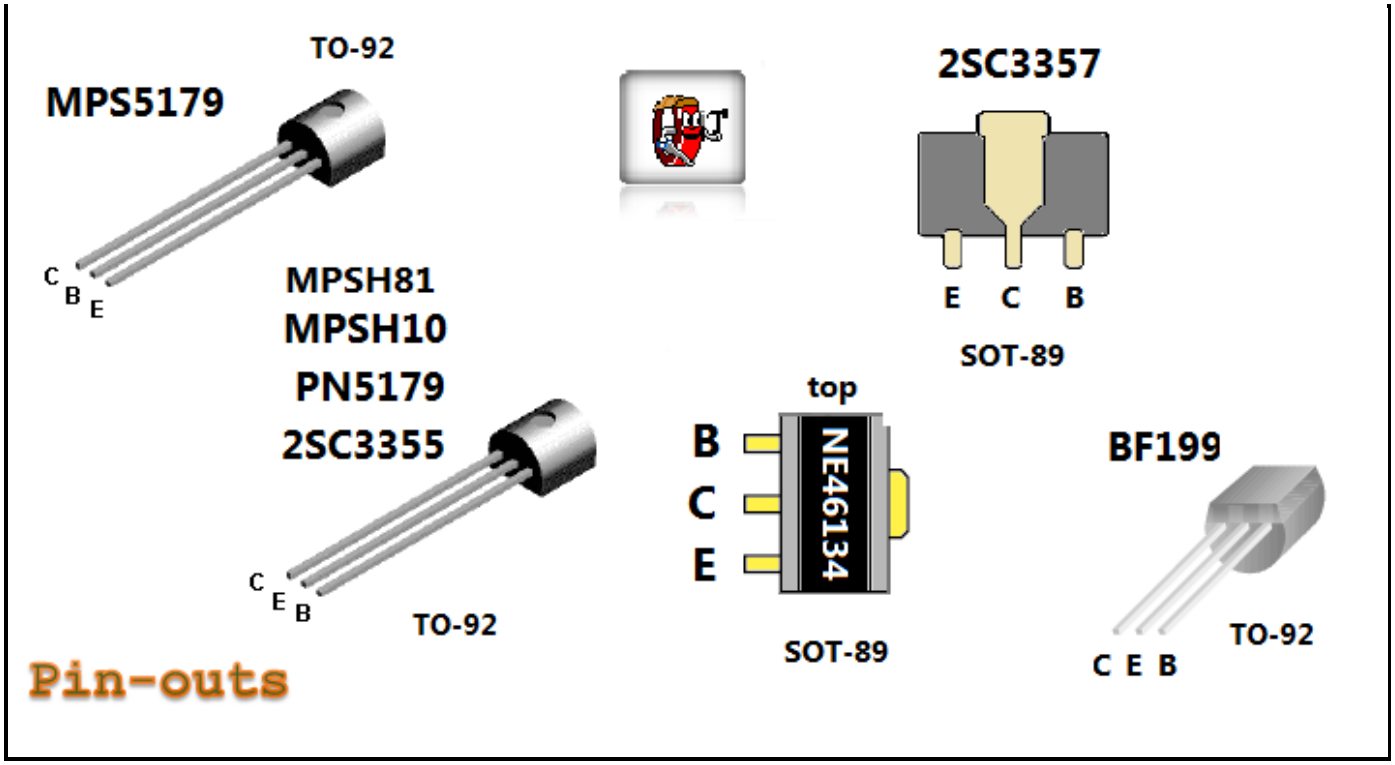
After trying a few circuits, I settled on a simple full-wave doubler featuring matched 1N4148 diodes driven by a JFET amplifier. Chapter 3 of Solid State Design for the Radio Amateur by Hayward and DeMaw for the ARRL = my key reference. Matched diodes (and a little luck), may suppress the 50 MHz signal up to 60 dB so only a single-tuned circuit follows the diodes.

The first JFET amp drives the diodes to improve harmonics + output voltage — an L-C-C Tee network matches the 2K7 Ω input to the 50 MHz signal generator output. I designed this matching network on the bench with the 2 diodes disconnected to avoid distortion during signal measurements. Using a 10X probe, I peaked the capactors for the greatest signal amplitude after finding the optimal L by educated trial and error. The inductor wound on a T37-6 was scrunched a little to further peak the L-C-C network.

I wound L1 with bare copper 26 AWG wire on a #8 bolt with coarse threads. To make enough space to solder on the 1/2 turn tap, I stretched the last turn with the other 3 turns still on the bolt to prevent stretching these coils. Click for a 'scope tracing at Point A. My 10X scope probe has ~ 15 pF capacitance and this affected the tuning — you can see some harmonics in the signal. With the 10X probe at Point B, I was nearly able to turn the circuit to 100.0 mHz, but still the 10:1 probe upsets the circuit somewhat.

This L - C tank tunes sharply and best with a non-conductive screw driver. I final tuned the L1 tank when the hycas amp was completed and connected to a 50 Ω terminated 'scope — eliminating the earlier tuning problems caused by the 10X probe. It seems that VHF requires more thought and care than HF when tuning resonators (tanks).

Point C is the maximum output (10.34 dBm) into a 50 ohm terminated 'scope. The 100.0 MHz signal is okay, but some builders might want to add low-pass filtering +/- an attenuator pad; or perhaps drive another feedback amp?


TO-92
MPS5179
C B E
MPSH81
MPSH10
PN5179
2SC3355
C E B
TO-92
2SC3357
E C B
SOT-89
top
B
C
E
NE46134
SOT-89
BF199
TO-92
C E B
Pin-outs

VHF to the Max — Miscellaneous 2012 Experiments

As a VHF newcomer, I need to make lots of circuits, measurements, mistakes and maybe — I might advance. My literature review revealed a big gap between popular, "for-fun", novelty-grade projects and the blinged-out circuits such as ultra low-noise LNA's featuring GaAs, MESFET, and pHEMT devices. Where do we find the middle ground projects? Likely on our own RF work benches.

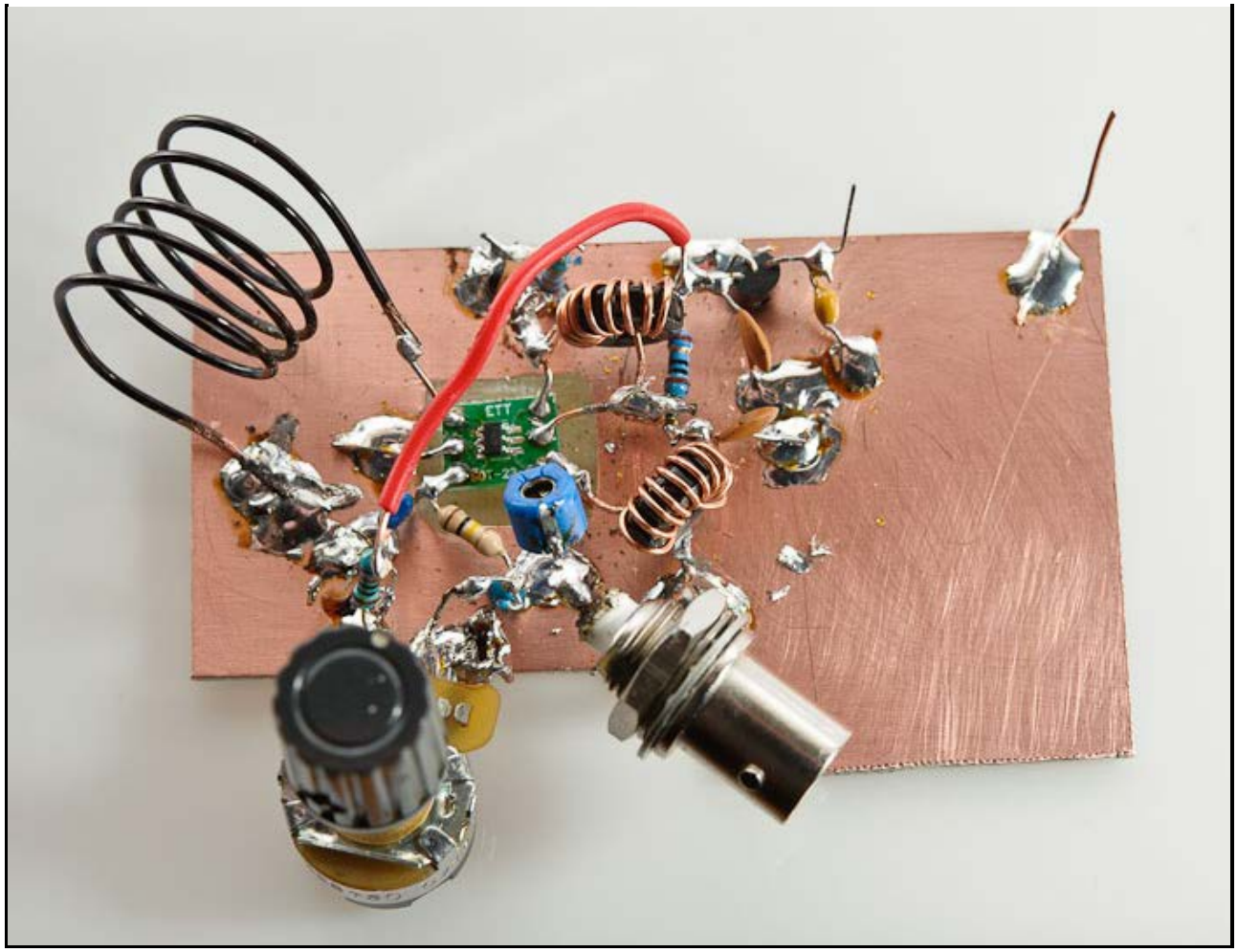
Numerous questions arose — Will Ugly Constuction work? Do I have the right test gear? Am I measuring the proper things? What about noise figure? All a bit overwhelming — but even improbable discoveries began as simple questions and observations. We solve far more difficult problems each day.

I'm learning that whether your a beginner, or a master, bench fundamentals rank supreme. Want to avoid oscillations in your high fT BJT VHF amplifier? Work towards high reverse isolation, strong S11 and S22, careful layout and employ techniques that suppress instability — nothing earth shattering huh? I'm told that acquiring the needed skills takes time and practice.

In early Winter 2011-2012, I built some VHF signal generators, amplifiers and a lots of junk. This web page documents a few of these experiments.

Section 1. Early MAX2606 VCO experiments
Section 2: A 2-band signal generator based upon the MAX2606 VCO
Section 3: 50 MHz VCO
Section 4: 50 MHz Receiver Pre-amp and Filter
Section 5: QRP — POSDATA: Z-Communications VCO Experiments

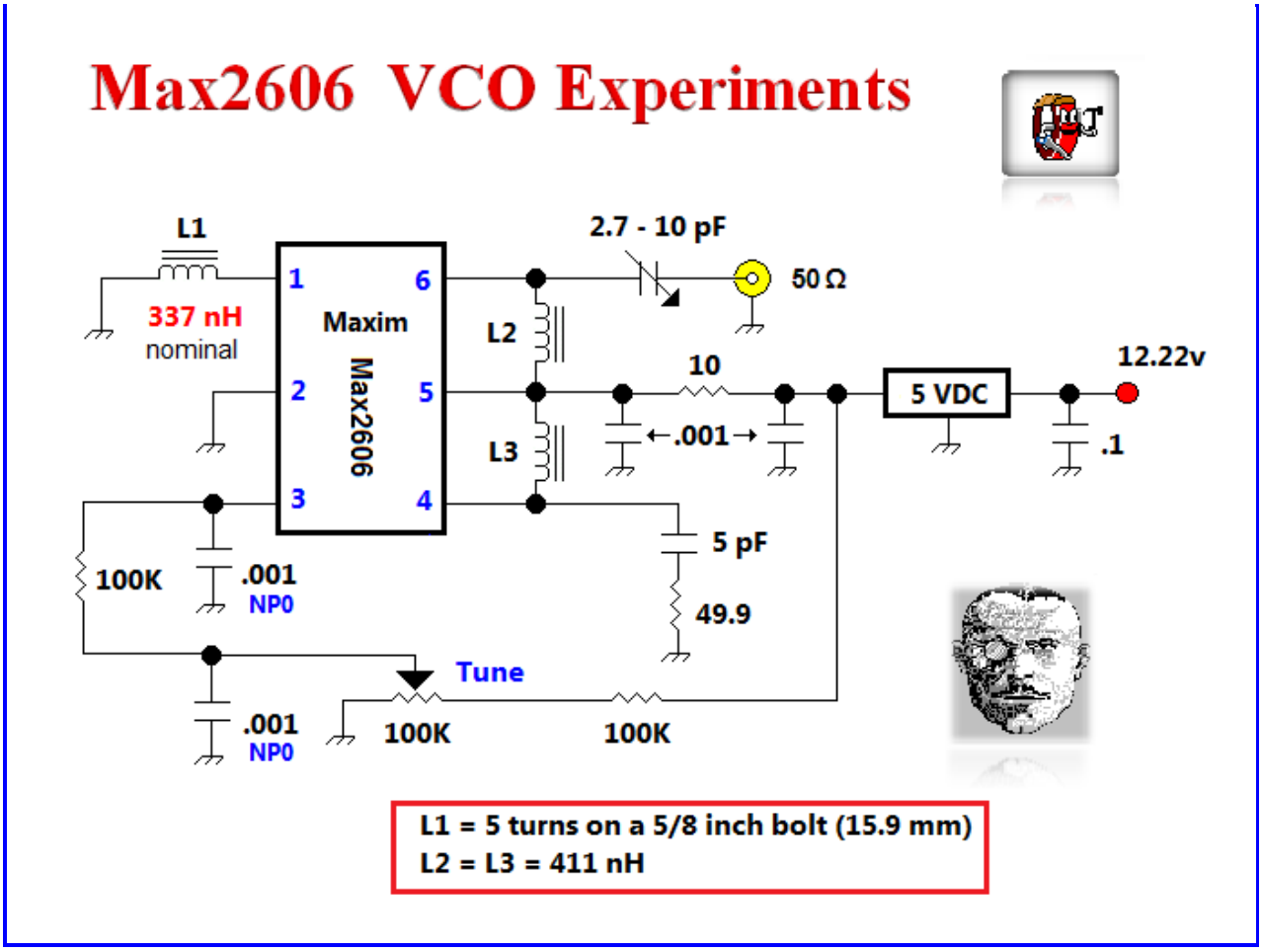
Section 1. Early MAX2606 Experiments



Above — My first of many MAX2606 VCO Experimental breadboards.

Maxim makes a series of cool SOT23-6 VCOs ranging from 45 to 650 MHz. Wanting a VCO to cover from ~100 to 106 MHz, I just had to try this chip. The datasheet provided all the online information I could find; save for a few homebrew, flea-powered FM transmitter projects that proved unhelpful.

After soldering the IC on a breakout (prototype) board using about 20X magnification, the proto-board was dropped onto a copper clad board with some of the copper ground away.



Above — My first MAX2606 VCO experiment. A coil lying on the workbench was soldered in as L1. I stretched it a little to set the lower band edge. Going from CCW to CW on the 100K tuning pot, I measured from 86.3 — 106.3 MHz. I'll write about L1 a little later, for making a good output network consumed my initial experiments.

The Maxim datasheet shows a simple output network consisting of 2 pull-up resistors with a maximum R of 1K. All the online FM transmitter designs I saw used 1K pull-up resistors and made no attempt at matching or even employing inductors in place of the resistors. I applied two 560 ohm pull-up resistors in place of L2 and L3 and the output looked distorted and low in amplitude. Click for a 50 Ω terminated 'scope tracing and click for a tracing with a 10X probe connected to a 10K load resistor. Unacceptable for even us scratch-builder RF experimenters.

Wanting to match the output into a 50 Ω load with high-pass L-network instead of just resistors, I calculated that an L of 250 to 800 nH, plus a capacitor from 2 to 10 pF might work, however, as a VHF newbie, my hopes weren't high. I went for a single-ended output and not knowing what to do, terminated the unused port with an AC-coupled 49.9 Ω resistor.

Choosing a 100 MHz test frequency (where I own a doubled 50 MHz xtal oscillator), I removed L1 to snuff the oscillator and measured an output return loss of ~ 17 dB with L2 = L3 = 298 nH and the variable cap set to 7.3 pF (I removed it and measured it). It took about 1 hour of trying different L values to get this return loss. I also learned that the L value and L to C ratio of L2 and its series capacitor greatly affects the purity of the output waveform.

Like all L-networks, the L and C *must* be correct to obtain a strong, clean output. Since C is variable, I tried various inductor values and further fine-tuned them by scrunching or expanding the L2 windings. *Compress or expand the L2 windings + trim the variable capacitor to peak the L-network at your desired frequency.*

From my experiments, the L3 inductance should be close to L2 for the best output voltage and symmetry, however, don't bother scrunching or expanding L3 because this won't greatly affect the output signal as long as the L3 inductance is close to L2.

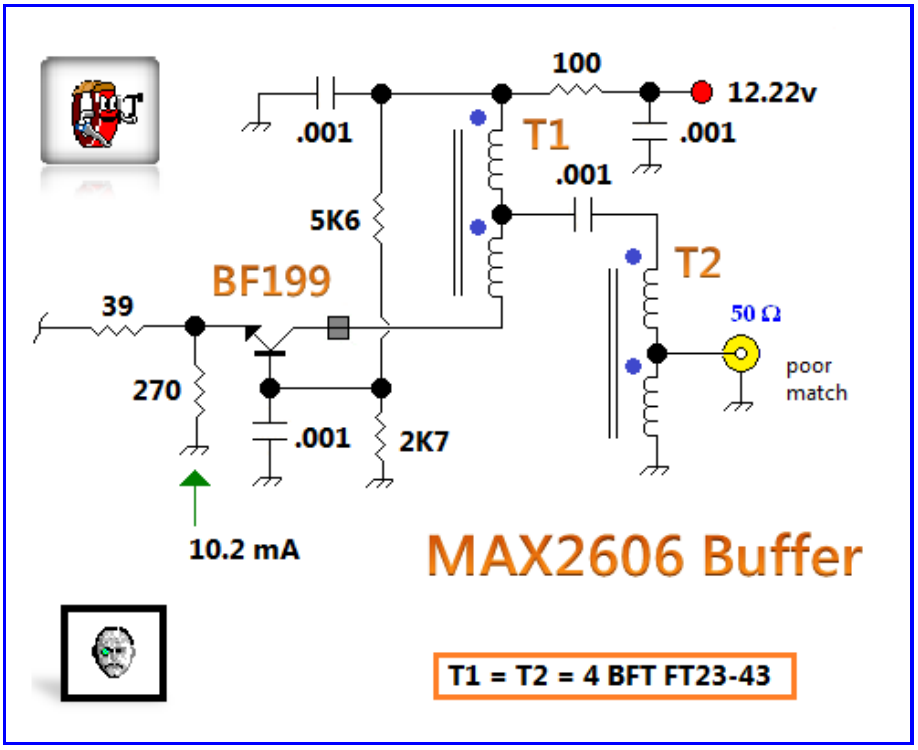
Click or click for some 'scope captures of poorly matched output L-networks — low gain and/or distortion appeared with mistuning.

Click for the output where L2 = L3 = 298 nH. I had the tuning set to 100.6 MHz for this screen capture, however, you can see the nice waveform and strong peak-peak voltage when compared to the pull-up resistor only versions.

The problem with a high Q L- network = low bandwidth. I wound L2 and L3 on T37-10 powdered iron toroids and later tried T37-6 toroids. The higher Q of the number 10 material seemed to translate into higher output voltage, but narrower bandwidth over the number 6 material toroidal inductors. Alternate breadboard. When tuning the MAX2606, signal amplitude changes with tuning frequency and a single L-network peaked somewhere near the middle of the tuning range further worsens this amplitude issue.

The *unbuffered* VCO waveform distorts somewhat as you move farther away from the peak L-network frequency. After trial and error, I settled in a compromise of L2 = L3 = 411 nH. This gave the best overall signal purity + amplitude from about 98 to 106 MHz with a set peak at ~ 102 MHz. Tuning below 98 MHz tended to really distort the waveform. Click for the *unbuffered* minimum and maximum frequency 'scope tracings.

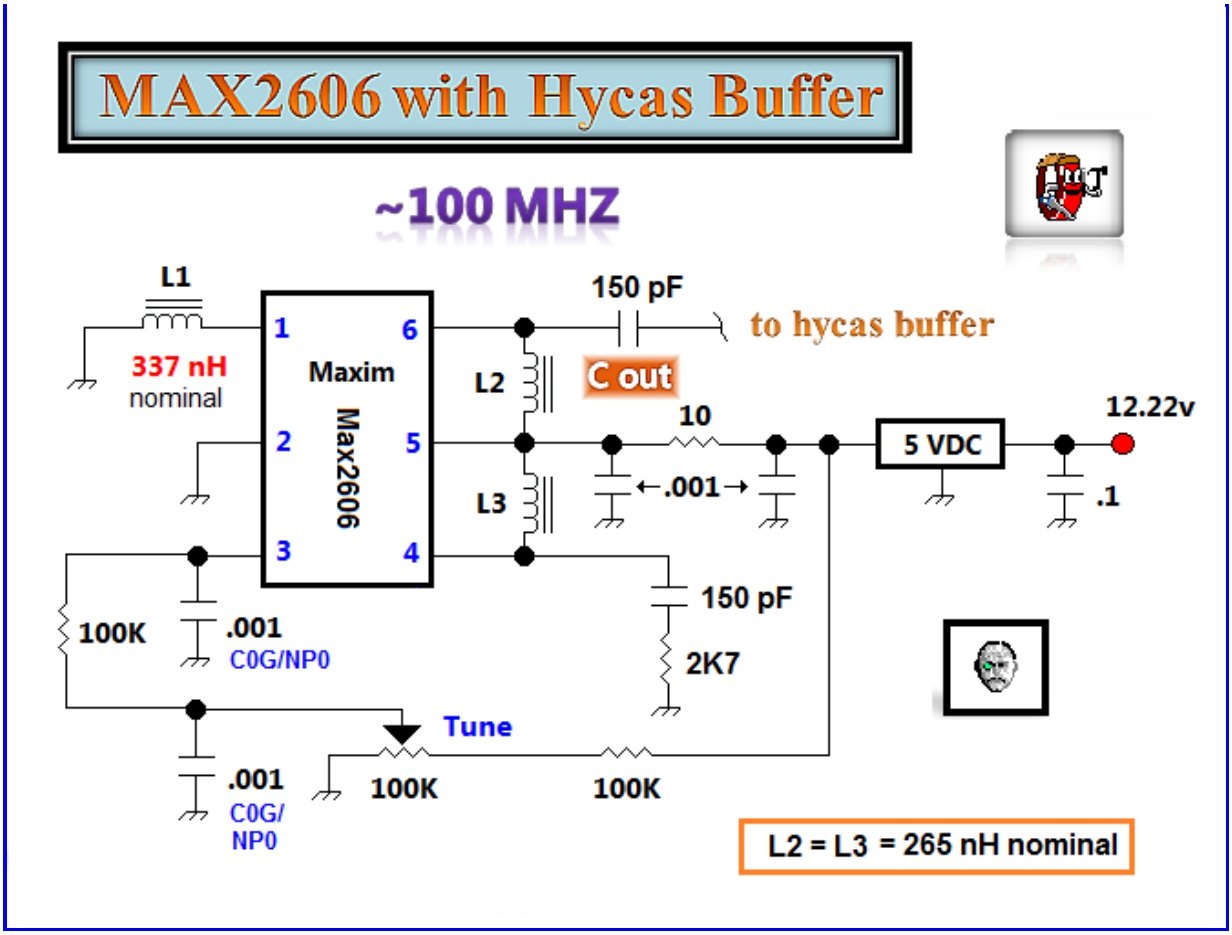
Placing an attenuator pad plus a buffer amp after the VCO dramatically reduced this distortion in my experiments, assuming the L-network components are correct and tuned.

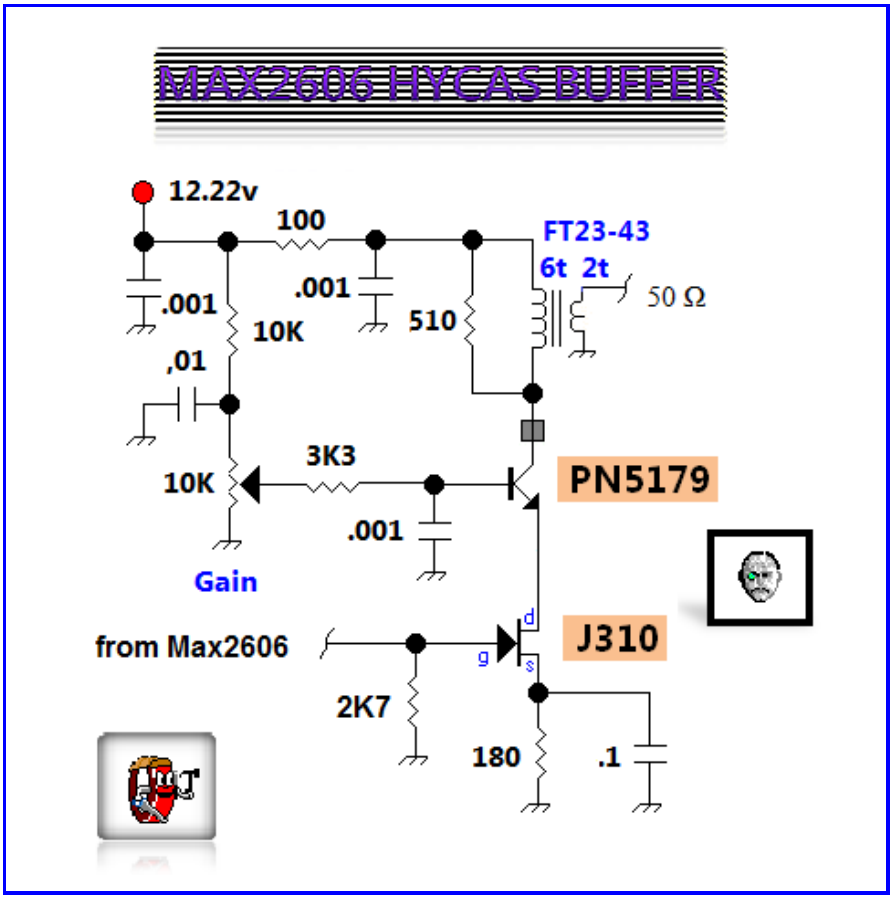


Above — A trial buffer first connected to output of the VCO schematic above. Two series-connected 2 transmission line transformers dropped the output impedance down for examination in my 50 Ω terminated scope — unfortunately the output return loss was poor (less than 10 Ω). Unlike the common base stage in a hybrid cascode amp, varying the bias to adjust gain worked poorly and lower applied bias caused distortion. I found it preferable to just fix the bias to get the greatest voltage gain + lowest harmonic distortion.

This buffer was discarded since the return loss was too low for my needs.

Testing a Hybrid-Cascode Buffer





Above — MAX2606 VCO into a hybrid cascode buffer at ~ 100 MHz. I also tried a buffer with a cascode of PN5179 transistors, however, the input Z of the buffer changed when adjusting the gain potentiometer and wrecked the input matching, so I stuck with the hycas buffer shown

above.

L1 and L2 = ~ 265 nH wound on T37-10 toroids. To peak the L-match at ~102 MHz, I originally placed a 100 pF in parallel with a 10 - 70 pF trimmer cap for *C out*. After tuning, I removed, then measured these 2 caps to discover that *C out* = ~150 pF; so I just substituted a 150 pF capacitor. The L1 windings were expanded a little to re-peak the output network — expanding the coils raises fCo, while compressing the windings lowers the cut-off frequency. I don't recommend omitting a trimmer capacitor unless your confident with your measurements.

Click for the maximum power 'scope tracing of the hycas amp with the L-match peaked at 102.2 MHz (3.44 dbM)

Once again, I terminated the secondary output of the MAX2606 with an load resistance equal to the main output with an approximately equivalent fixed-value capactor — I'm not sure if it's needed, but it works okay and I stuck with it.

Up at ~ 144 MHz

Click for another experiment with the fore mentioned MAX2606 into the hycas buffer shown above at ~ 144 MHz. I peaked the L-match for 142.2 MHz and the output power = a surprising 7 dBm. The output is a clean sine wave from 135 to 169 MHz, although the signal amplitude varies widely. I also peaked it at ~144 MHz. I wound L1 - L3 with #21 AWG on a #10 bolt. Photo 1 Photo 2.

Let's build something useful with the MAX2606 based upon the experiments thus far...

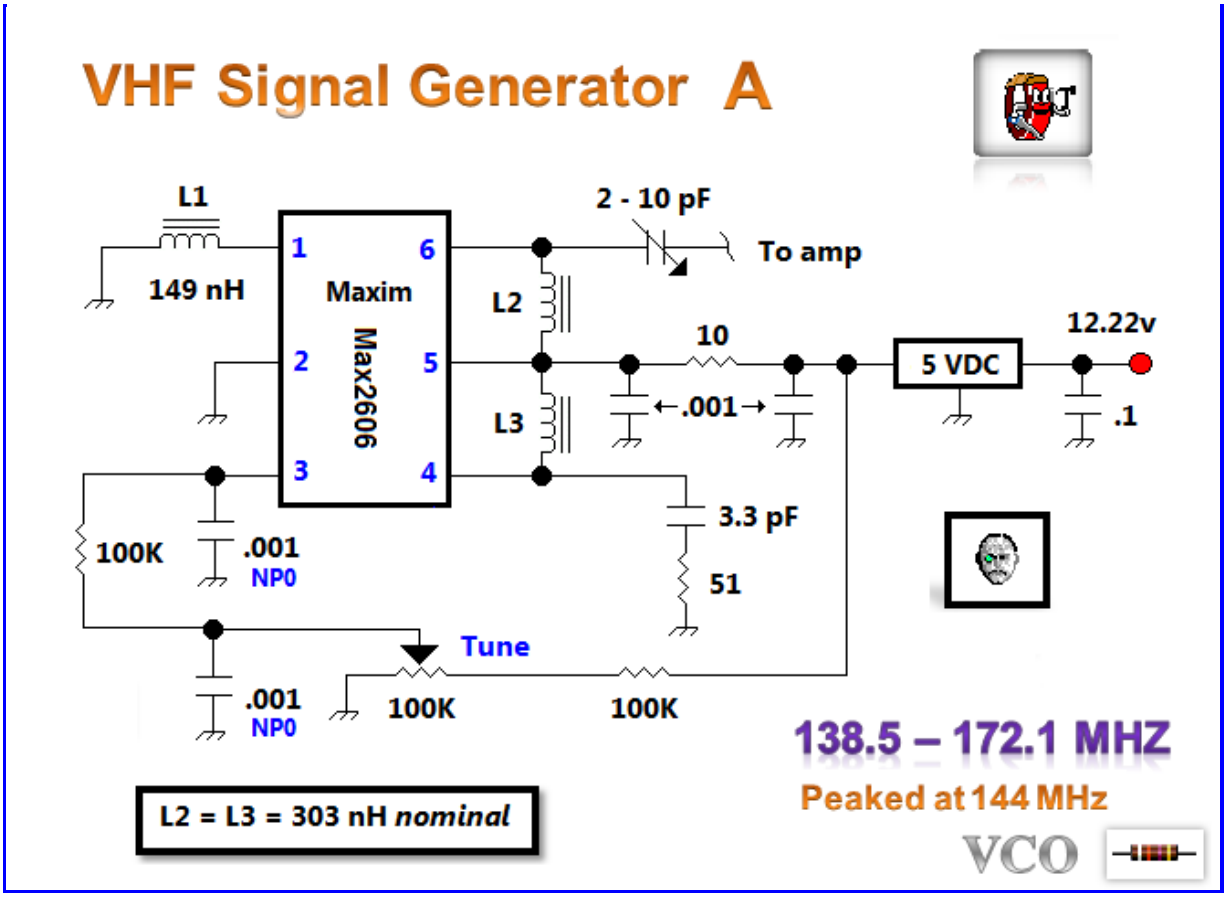
Section 2. A 2-band Signal Generator Based Upon the MAX2606 VCO

Above — My dual VCO based on the MAX2606. This general-purpose signal generator will start my VHF circuit development in these 2 frequency bands. The importance of owning good signal generators can't be overstated — while not engineer grade, this box features a clean sine wave, strong output return loss and 1 KHz or better tuning resolution.

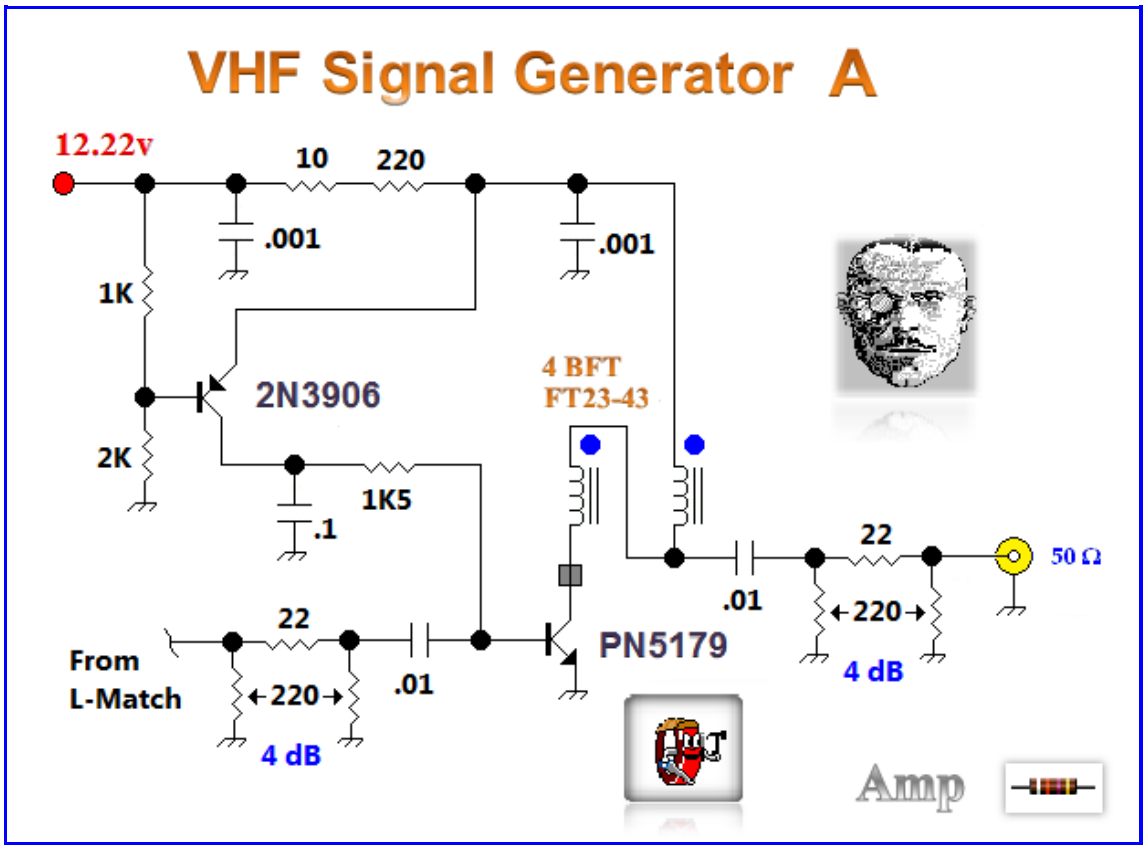
The center ON-OFF-ON toggle switch only turns on 1 oscillator at a time. The top 2 (ten-turn) tuning potentiometers lack knobs *(I'll get some later)* and the black knobs below them are attached to *unused* potentiometers. Initially, I planned to employ front panel gain control and drilled holes and fitted 10K pots in the chassis — I later decided to control the output power with outboard variable attenuators and skipped front panel power control for simplicity sake. Click for a side view of the unfinished project. Click for an early photo of the VCO **A** breadboard — I attached a shield to the copper clad board to help isolate the 2 VCOs.

I chose the frequency band **A** (138.5 - 172 MHz) to include the Ham 2 meter band + local commercial/service VHF segment and band **B** (98.919 - 109.06 MHz) to capture the FM broadcast band above 98 MHz.

Band A: 138.5 - 172 MHz



Above — The VCO schematic **A**. An L-match peaks the output at 144 MHz into 50 Ω. The beauty of this circuit = simplicity; just 2 active devices give low distortion and a strong output return loss (S22) on the 2 Meter Ham band.



Above — The output amplifier for VCO "A". Click for a schematic with some analysis at 144 MHz (the frequency I'll use the most). I spent a lot of time trying to develop a 50 Ω output Z voltage amp up at ~ 144 MHz. My attempts to employ shunt and series feedback gave generally poor results — stray reactances plague the standard FBA designs that work great under ~70 MHz.

Since the MAX2606 has an output L-Match, getting a high output return was my only goal — grounding the PN5179 emitter, employing a 4:1 transmission line transformer and biasing for ~15 mA emitter current did the trick. A 4 dB pad on the NPN input further establishes a strong S22 (output return loss).

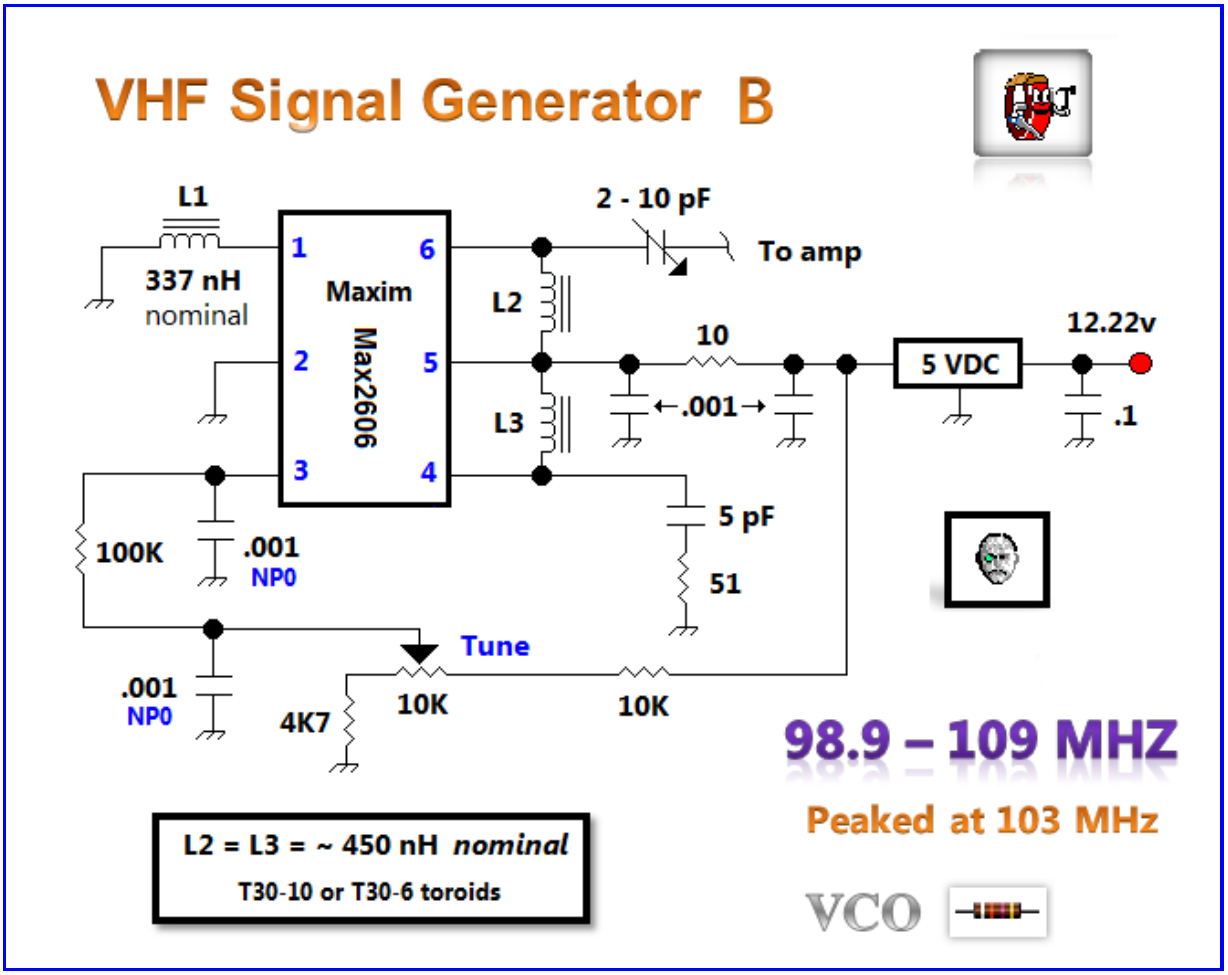
As possible, I attach additional 50 Ω outboard attenuators on my signal generators, however, the S22 on the stock generator should be okay across the VCO range due to the two 4 dB pads.

I measured the S11 on a prototype NPN amplifier by using the MAX2606 VCO shown to drive a MMIC with a 16 dB 50 Ω attenuator pad on the output and connecting it to the RF port of my return loss bridge. Thus, the actual VCO helped me design the final amplifier which buffers it in my final build. I kept the PN5179 and all other leads short as possible.

The output power looks like a sine wave when tuned from CCW to CW: With the 10-turn tuning potentiometer set to CCW [138.5 MHz], the output power ranges from -1.9 dBm; hits a peak of 0 dBm at 144 MHz and then gradually drops to -11.2 dBm at CW [172.1 MHz].

I moved away from the hycas buffer amp to simply my design and increase reproducibility. A MMIC might also work well, however the high current drain, potential for instability and biasing considerations introduce new problems — a simple wrap around PNP-biased NPN amplifier works okay. Sometimes the best solution = the simplest.

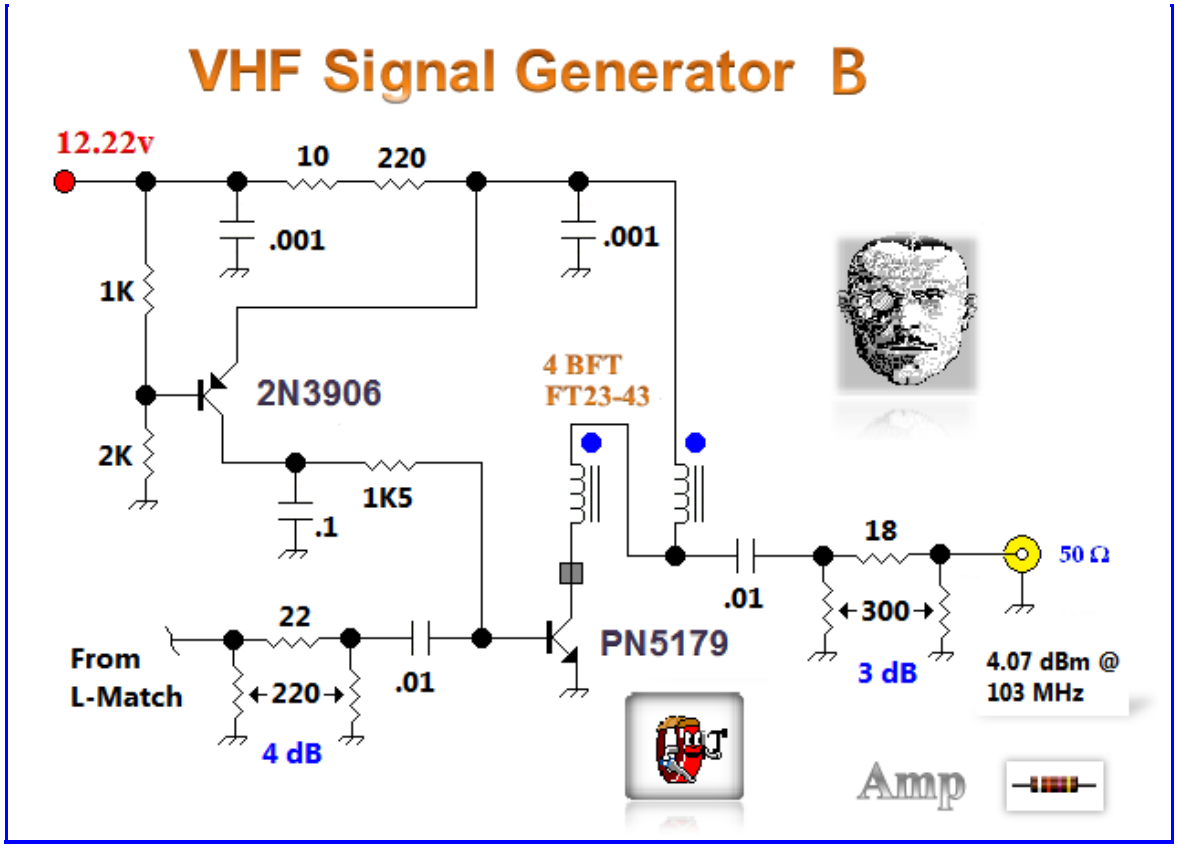
Band B: 98.919 - 109.06 MHz



Above — The VCO schematic **B**. An L-match peaks the output at 103 Mhz, although compressing or expanding L2 and tweaking the 2 -10 pF capacitor can peak the L-network anywhere in the tuning range. I limited the tuning range by adding a 4K7 R to the tuning pot to enable better matching and fine tuning.

In all cases L1 = an air inductor wound on a coarsely threaded bolt using 20 or 21 guage enamel-coated magnet wire for mechanical stability. I set the lower band edge of the VCO by setting the tuning pot to CCW and compressing or expanding the L1 inductor to get the frequency shown.

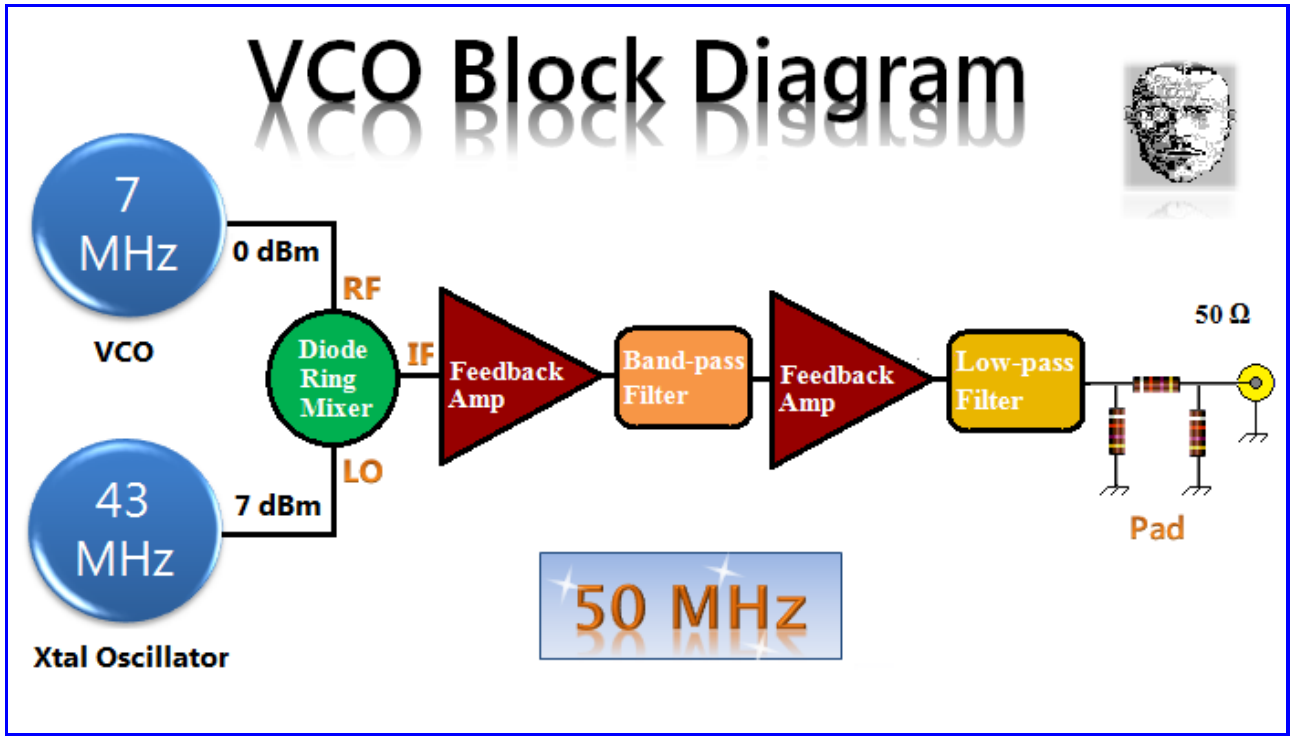
For L1, I found that excessively long leads can create unwanted oscillations and my coils are just a few mm above the copper board. Some builders properly mount their coils in a upright "smokestack" fashion and/or well away from the nearby metallic chassis or copper PCB to minimize Q losses + inductance changes. My coils lie well away from the metal chassis walls.



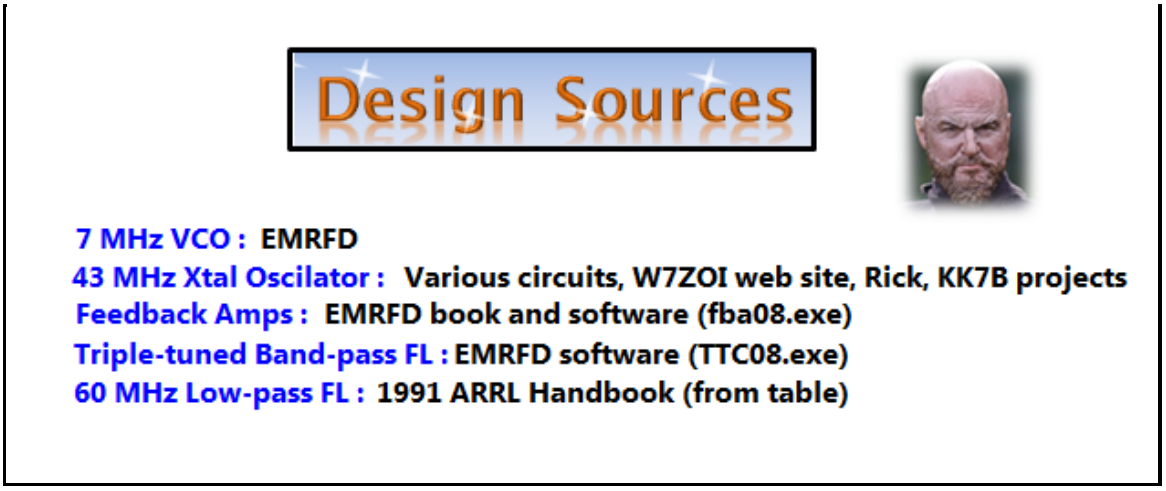
Above — The VCO **B** buffer/amplifier schematic. A clean sine wave appears across the entire tuning range — output return loss tuning from minimum to maximum was > 21 dB before I added the 3 dB pad.

Click for a photo of the completed, partially labelled project during the final tune up with all the boards bolted in.

Section 3. 50 MHz VCO

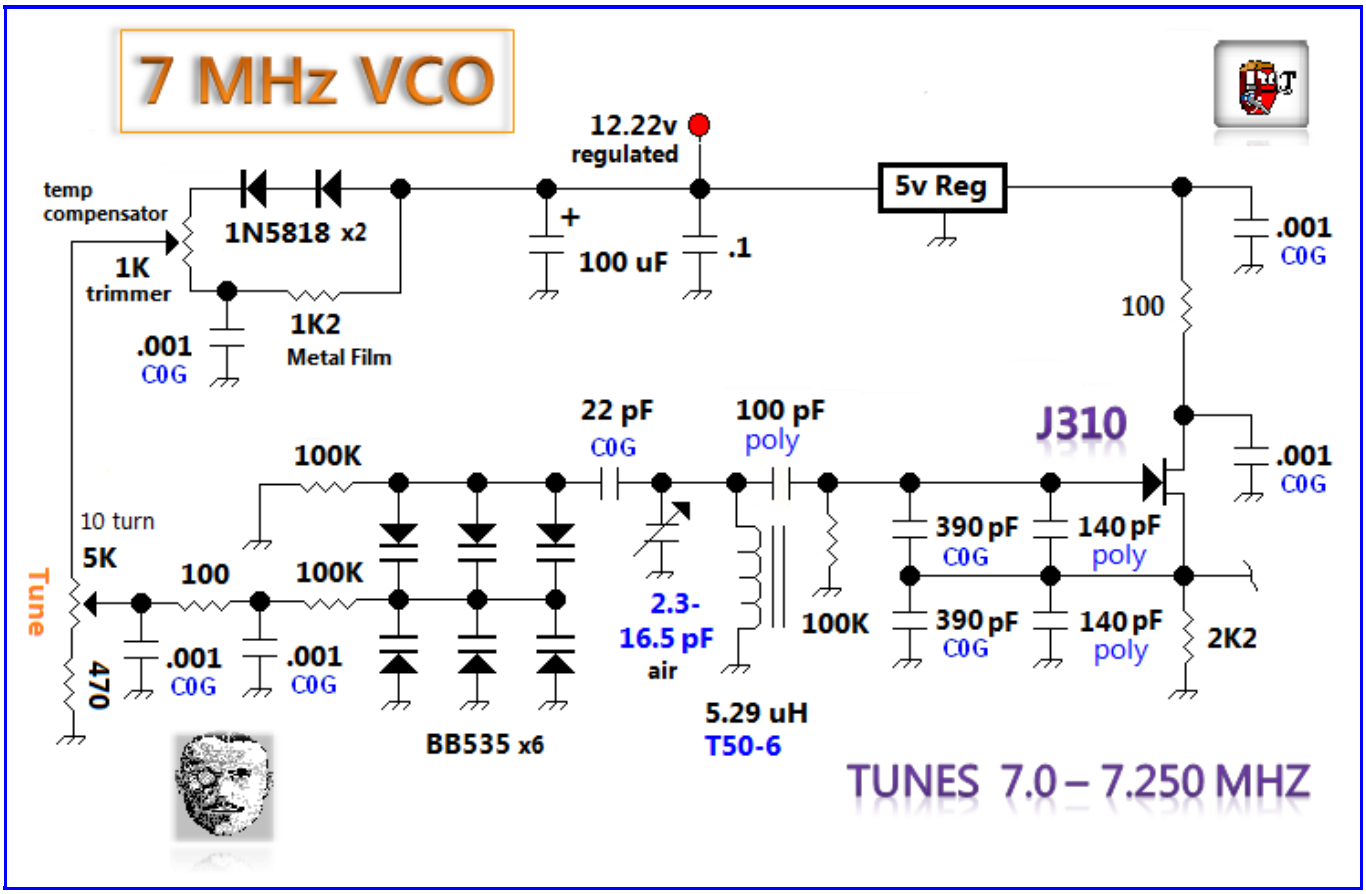


Above — Block diagram of the 50 MHz VCO I designed and built in February-March 2012. Click for a photo.



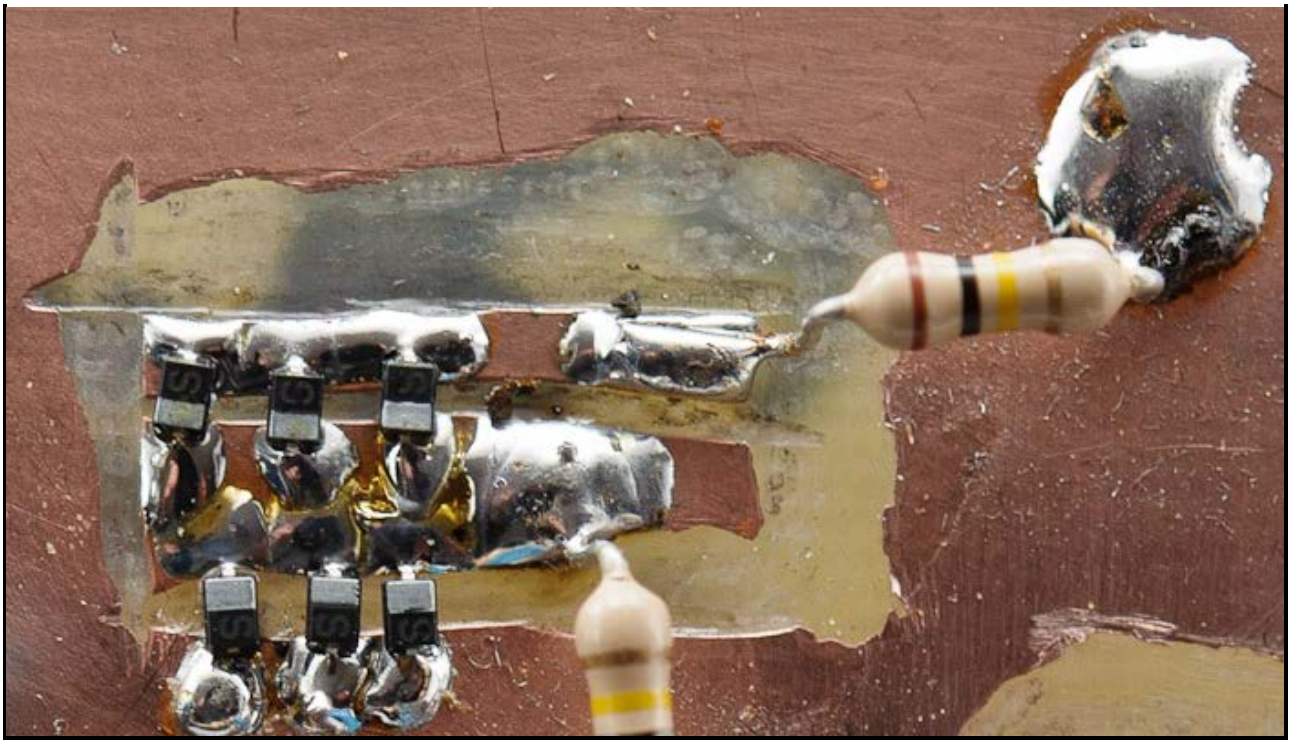
Above — Like most of you, I'm just an amateur designer who relies on others for example circuits, design procedures and inspiration. These cited references plus hard work drove my experiments. This project succeeds the Miscellaneous RF Experiments web page from 2011 — QRP SWL HomeBuilder evolves as I do.

7 MHz VCO + Buffer Amplifier [0 dBM output power]



Above — 7 MHz Colpitts VCO schematic.

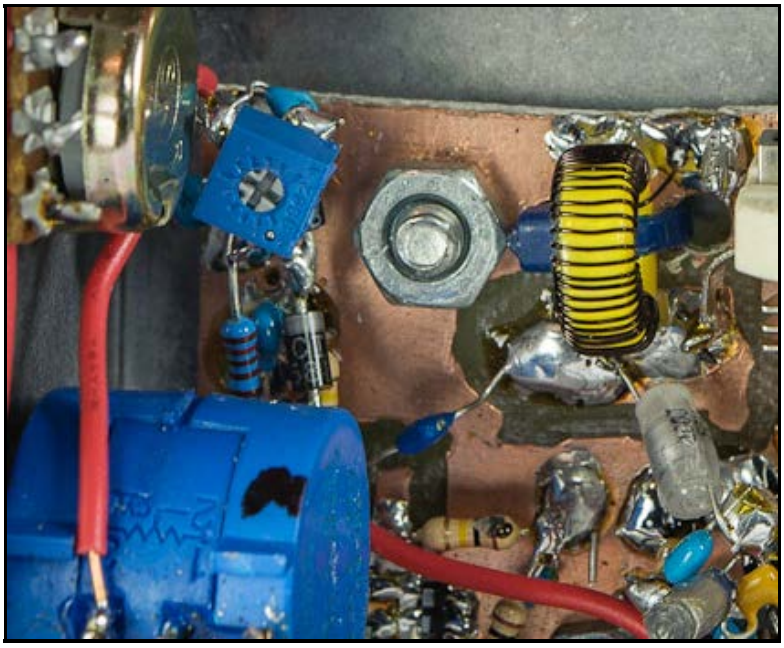
This VCO tunes ~7.00 to 7.250 MHz, although a wider tuning range occurs if you allow the tuning diodes to drop to 0 VDC with the 5K tuning pot cranked CCW. A 470 ohm resistor keeps about 1 VDC on the varactors at the lowest tuning frequency/applied reverse DC voltage.



Above — A macro photograph of the six BB535 varactors soldered on the VCO breadboard. With 0 applied reverse DC voltage, their total C = 43.5 pF. I left room for up to 4 more diodes, but didn't need them.

Macro photography provides an excellent way to inspect SMT parts — apart from all the fiberglass dust on the board, no shorts or other problems arose when soldering. Next to pF-value chip capacitors, these SMD varactors proved the most difficult surface-mount parts I've breadboarded to-date. Using clear tape, I tape my SMT parts to the PC board when soldering. With tape, you can still make tiny device placement adjustments with a pick or tweezers and yet the device holds steady enough to solder. I recently obtained a microscope for SMT work, although didn't need it for these diodes.

Striving for lower phase noise meant properly applying high Q tank parts — I soldered in 3 pairs of high-grade BB535 varactors and arranged them anti-parallel to avoid forward conduction + even harmonics. I also limited the AC voltage swing they "see" by connecting them to the L with a 22 pF capacitor. Tight windings of #28 gauge magnet wire on a T50-6 toroid formed the inductor. 4 number 8 bolts anchor the 7 MHz VCO board to the chassis and prevent board warp + movement.

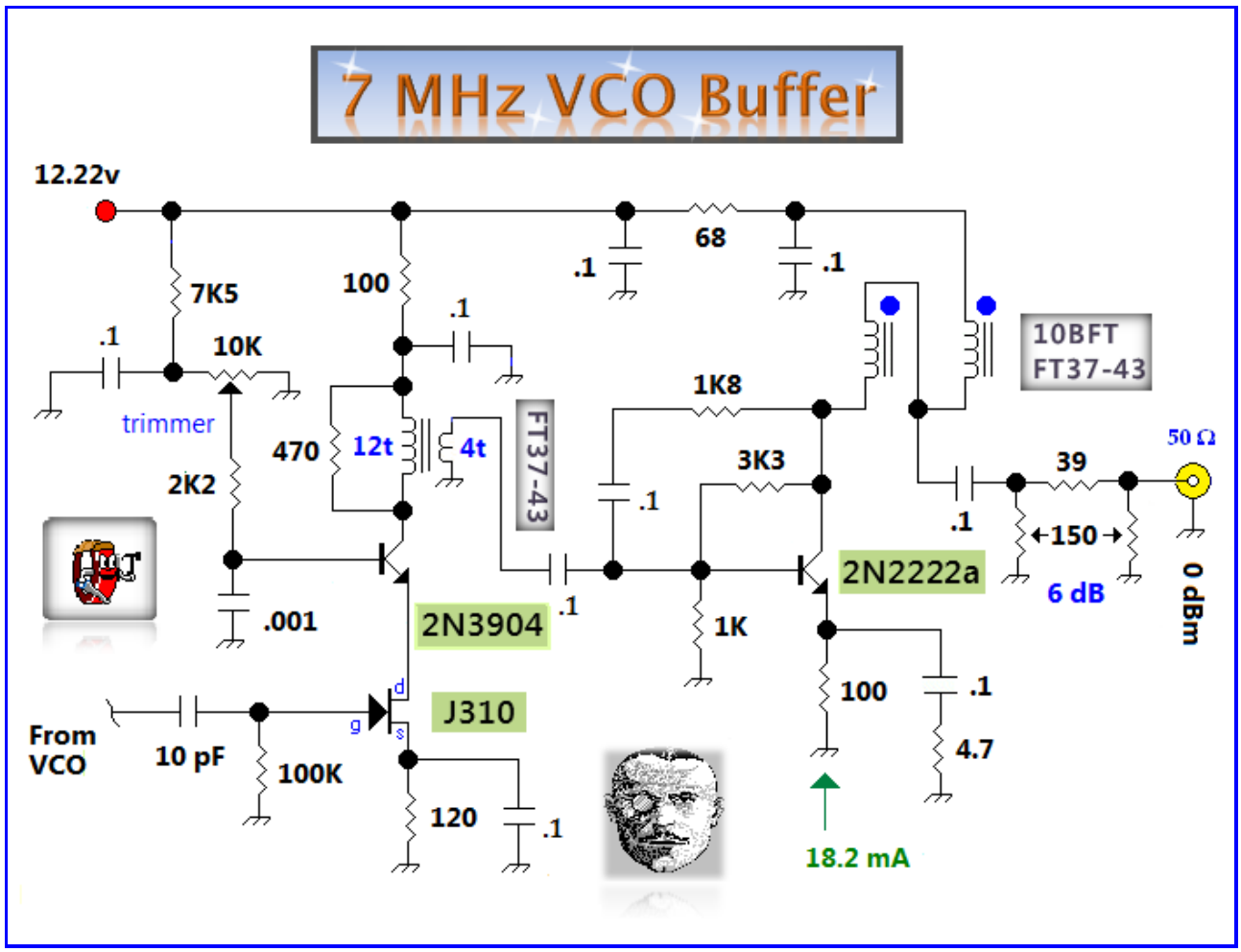


Above — a view of the square blue 1K temperature compensation (tempco) trimmer potentiometer.

To aid temperature compensation, I included 3 polystyrene capacitors in the base VCO — the tempco circuitry represents about 16 hours of work from December 2011. Click or click for photos of the bread board before the tempco parts were soldered on — the temporatry BNC connector lies in the background was removed after testing.

With care and patience my lid-on 1 hour temperature drift = ~10 Hz. My temperature compensation strategy worked because I took the time to measure and then determine how to cancel temperature drift in this 1 circuit — your results will vary and *experimentation* remains the key to temperature compensating VCOs and VFOs.

See EMRFD, and the VFO-2011 + QRP Modules 2011 web pages for more tempco information.

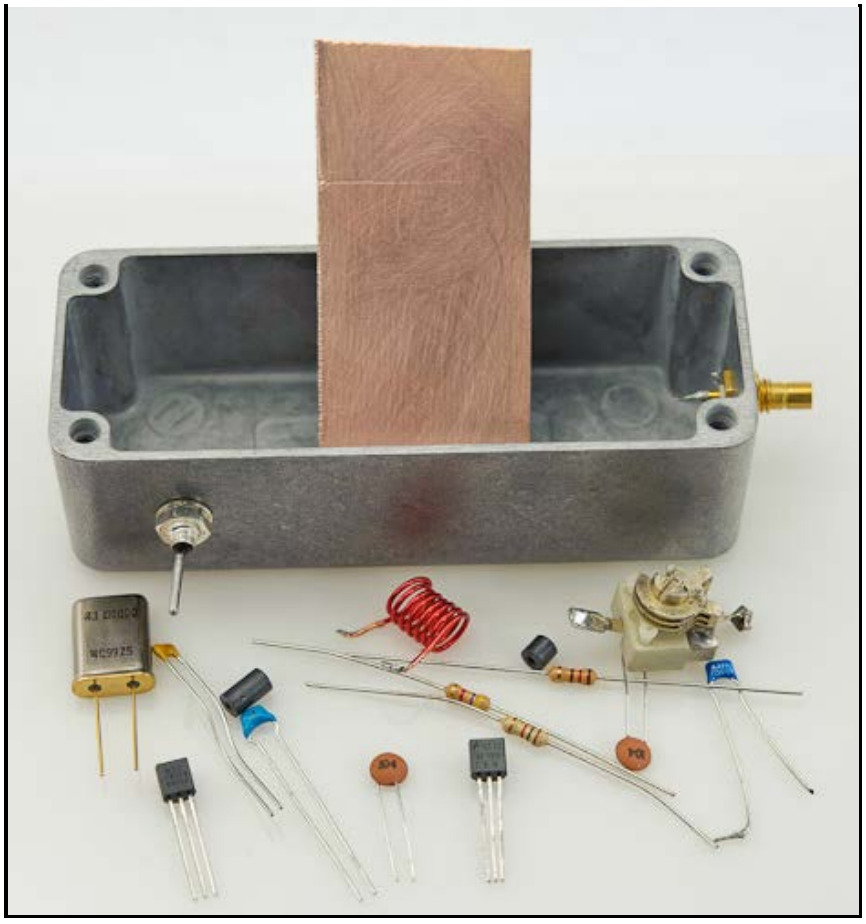


Above — 7 MHz VCO buffer/amplifier. I adjusted the 10K trimmer pot on the hycas buffer for exactly 0 dBm drive. Even before adding the 6 dB attenuator pad, the output return loss = 23.8 dB @ 7.0 MHz.

Originally, I wanted a VCO output of 7 dBm and applied ~18 mA emitter current in the final amp to preserve signal fidelity and eliminate the need for a low-pass filter. This buffer works great up to an output power of ~10 dBm: above 10 dBm or so, distortion occurs and you'll need to add a low-pass filter.

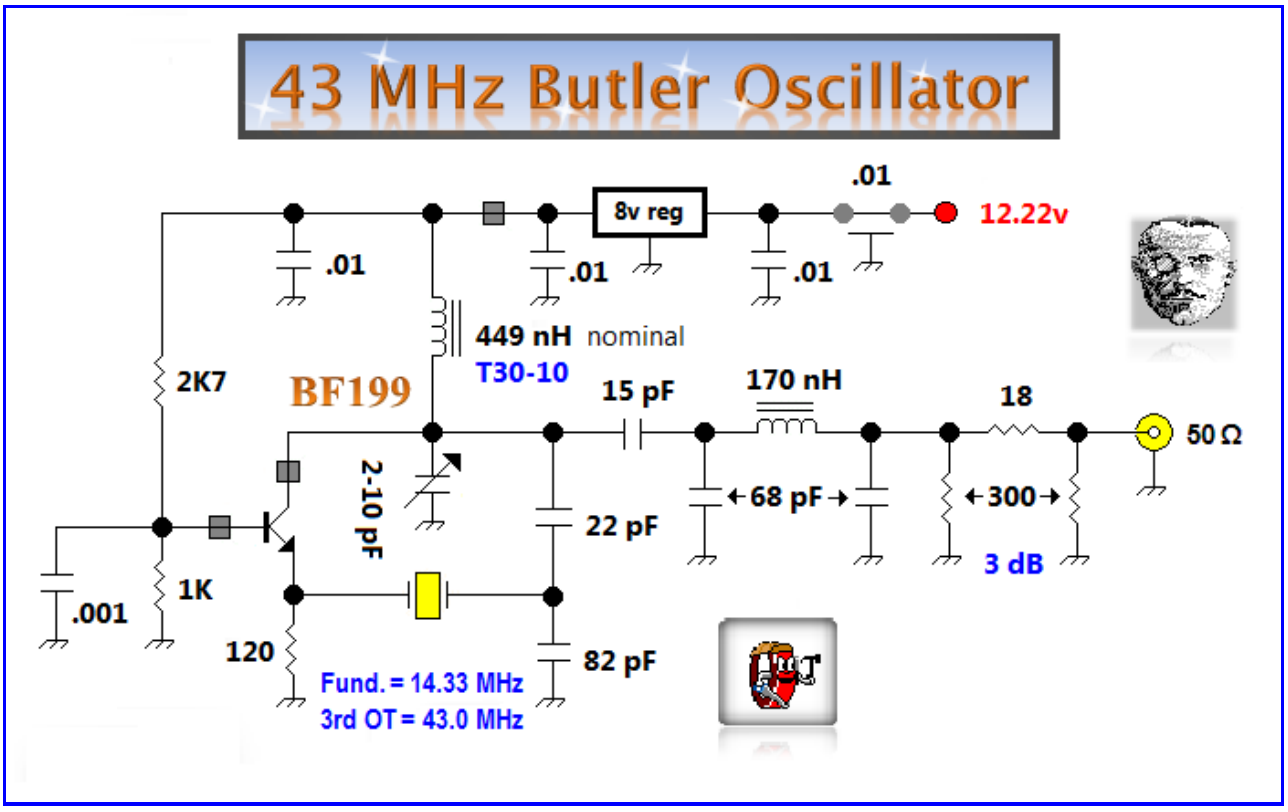
Adjust the hycas trimmer pot for whatever output power you seek, but If you're ever using this buffer for 7 -10 dBm output power, drop the 6 dB attenuator pad to 3 dB. This drops the drive level to maintain low harmonic distortion (2nd harmonic down > 35 dBc).

43 MHz Butler Xtal Oscillator [6.4 dBm output power]



Above — Some Butler crystal oscillator parts prior to the build. Since this Butler will go inside a box containing a VCO and some high gain amplifiers, it would be foolish to not stick it in an RF-tight box. On the Hammond chassis above, you'll notice a feedthrough capacitor for the 12 VDC and an gold colored SMA connector for the output.

Since the required L = over 400 nH, I opted for a toroidal inductor wound on a T30-10 instead of the air coil shown in the photo.

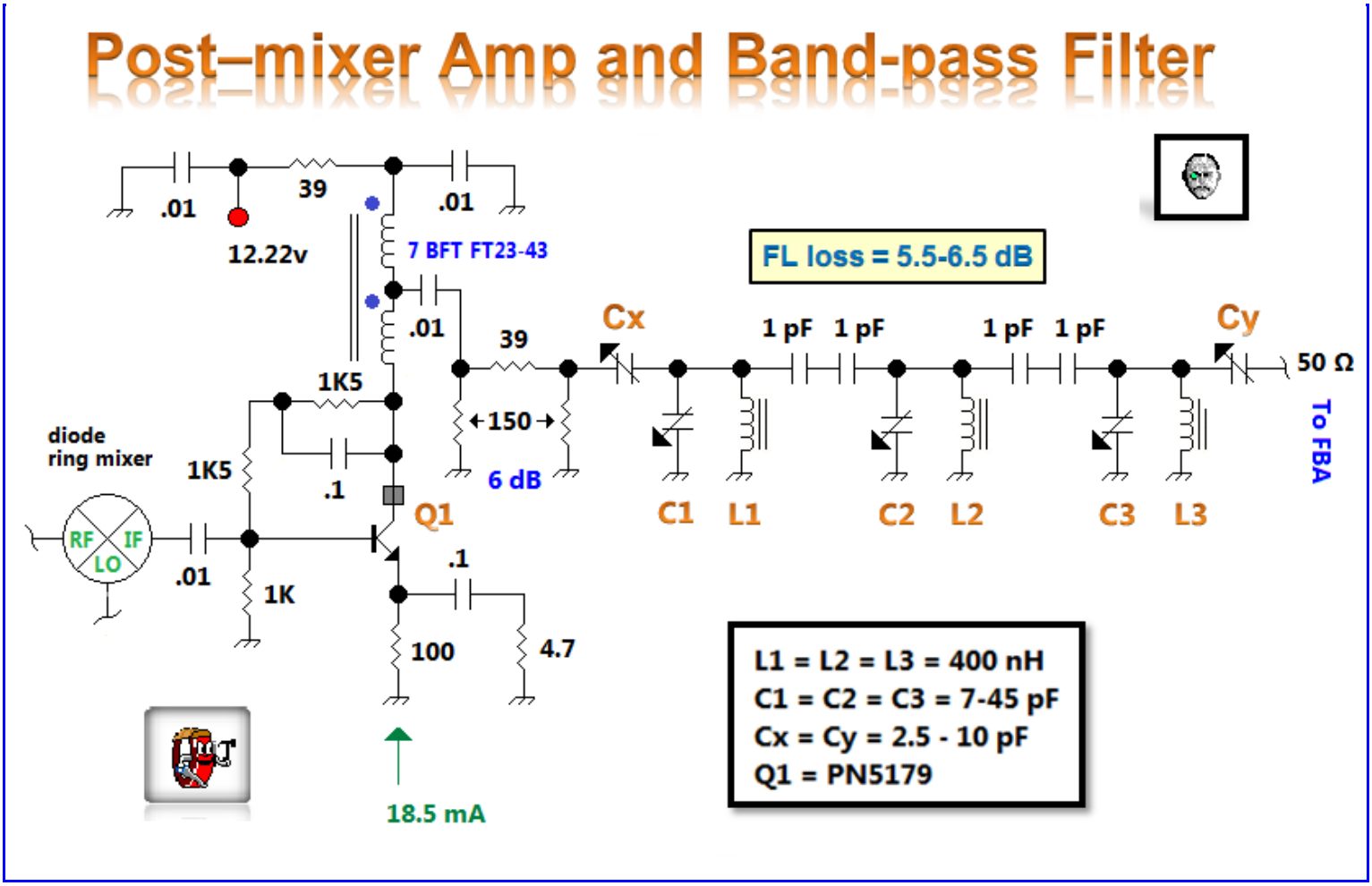


Above — 43 MHz Butler overtone oscillator schematic. The highest power I could muster = 6.4 dBm (close enough to 7 dBM).

This Butler looks good on FFT. Click, Click or click for 3 'scope captures. Despite trying to milk maximal power, the 2nd harmonic is over 40 dBc down. Click for a snap shot of the completed oscillator.

Above — The original Butler oscillator before adding the pi low-pass filter. The bolt (seen at top right) will also pass through the outer VCO chassis to hold this sub-chassis in place. Click for a bigger photograph.

Post-Mixer Amplifer and Triple-Tuned Filter



Above — Schematic of the diode ring mixer, Q1 feedback amp and the triple-tuned filter. *I used a MCL SBL-1 mixer*. L1 - L3 were wound on T30-10 toroids. I bought my #10 and some #12 toroids from the great folks at Debco Electronics.

The post-mixer feedback amp data at 50.0 MHz (isolated from the mixer and pad + filter) : Emitter current = 18.5 mA, S21 =18 dB, S11 = -24.4 dB, S22 = - 21.5 dB. *(S21 = power gain; S11 = negative of the input return loss; S22 = negative of the output return loss).*

The 6 dB pad helps absorb signal reflections from the filter caused from stray reactance plus capacitance variations caused by coupling the 2 tanks with only 0.5 pF (2 series 1 pF capacitors with a +/- 0.25 pF tolerance!)

Preliminary filter alignment: Peak your filter however you want — but here's how I peaked my filter with a crystal controlled 50.0 MHz signal generator connected via a temporary BNC connector tack soldered onto the copper board and wired to the Q1 input. Terminate the filter with a ~50 Ω resistor, or a temporary BNC connector plugged with a 50 Ω resistive terminator.

Connect the signal generator to Q1 and peak C1, C2 and C3 (in that order) using a 10X 'scope probe. It's better measure with your probe at point C2 when tuning C1 since this reduces mistuning caused by probe capacitance — measure at point C3 when tuning C2 etc.

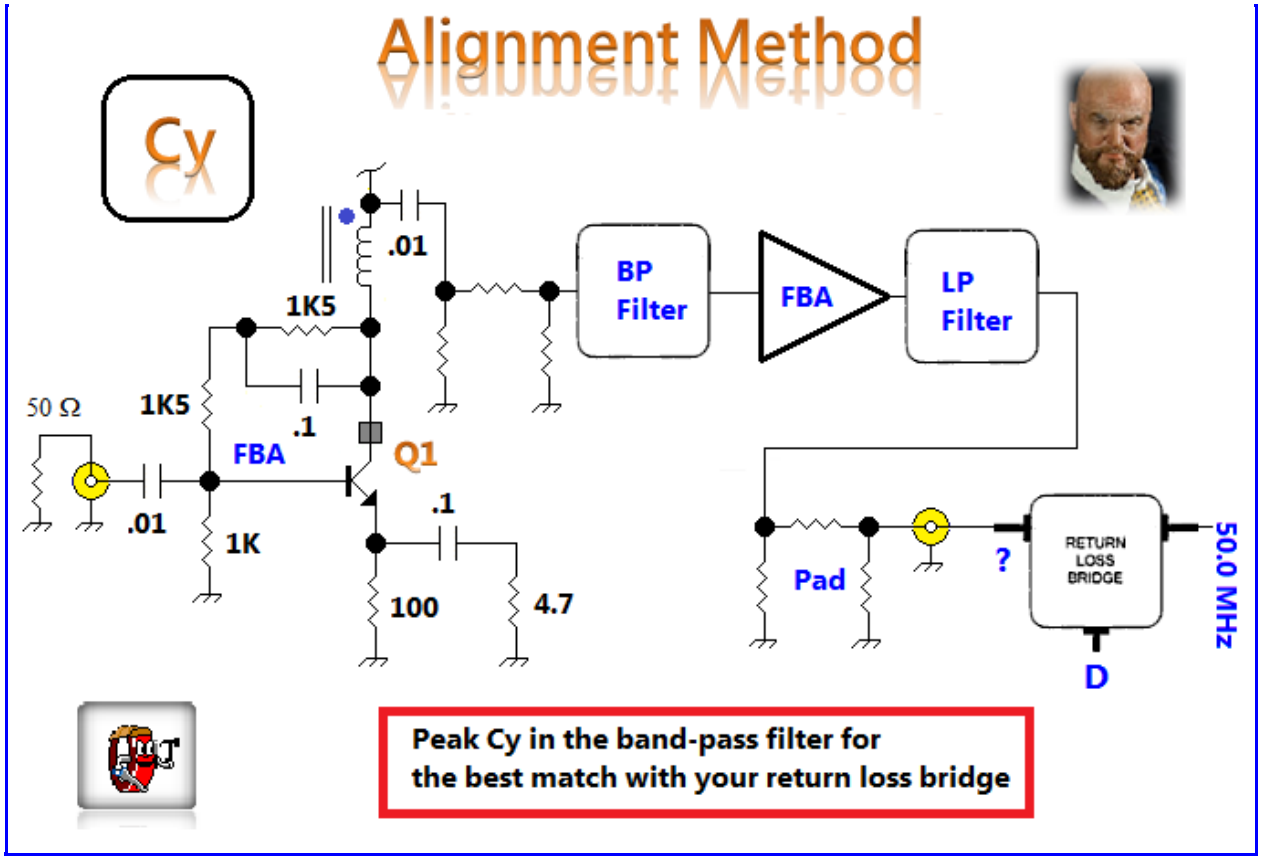
Then peak C3, Cx and Cy with the probe touching the terminating 50 Ω resistor. *It easier to perform the first tune-up with a 10X probe going sequentially from C1 to C3 since these peaking capacitors tune pin sharp.*

After the preliminary tune-up, if possible, connect a temporary BNC connector to the output and re-peak all the caps with a 50 Ω terminated scope; this boosts sensitivity and eliminates 'scope probe capacitance.

Perform pentultimate 50.0 Mhz alignment after you add the post-filter amp, low-pass filter and the 3 dB pad. Capacitor Cy critically sets the output return loss of Q2 and when properly matched, establishes a 50 Ω termination for the triple-tuned band-pass filter.

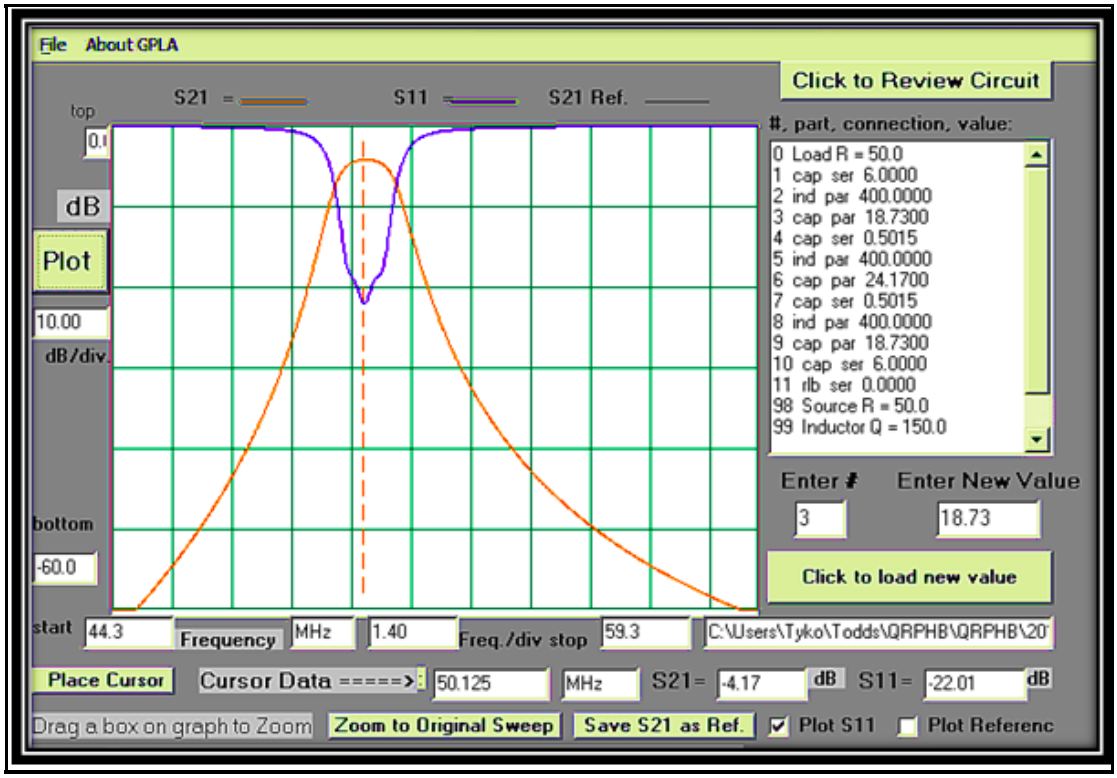
You can also match Cx by connecting a return loss bridge to the input of Q1 and terminating the RF chain with a 50 Ω resistor, although tuning Cx only changes the input S11 a little. In 1 experiment, I replaced Cx with 6 pF and it worked okay.

I wonder how I ever managed before making a return loss bridge: the workhorse of the QRP workbench.



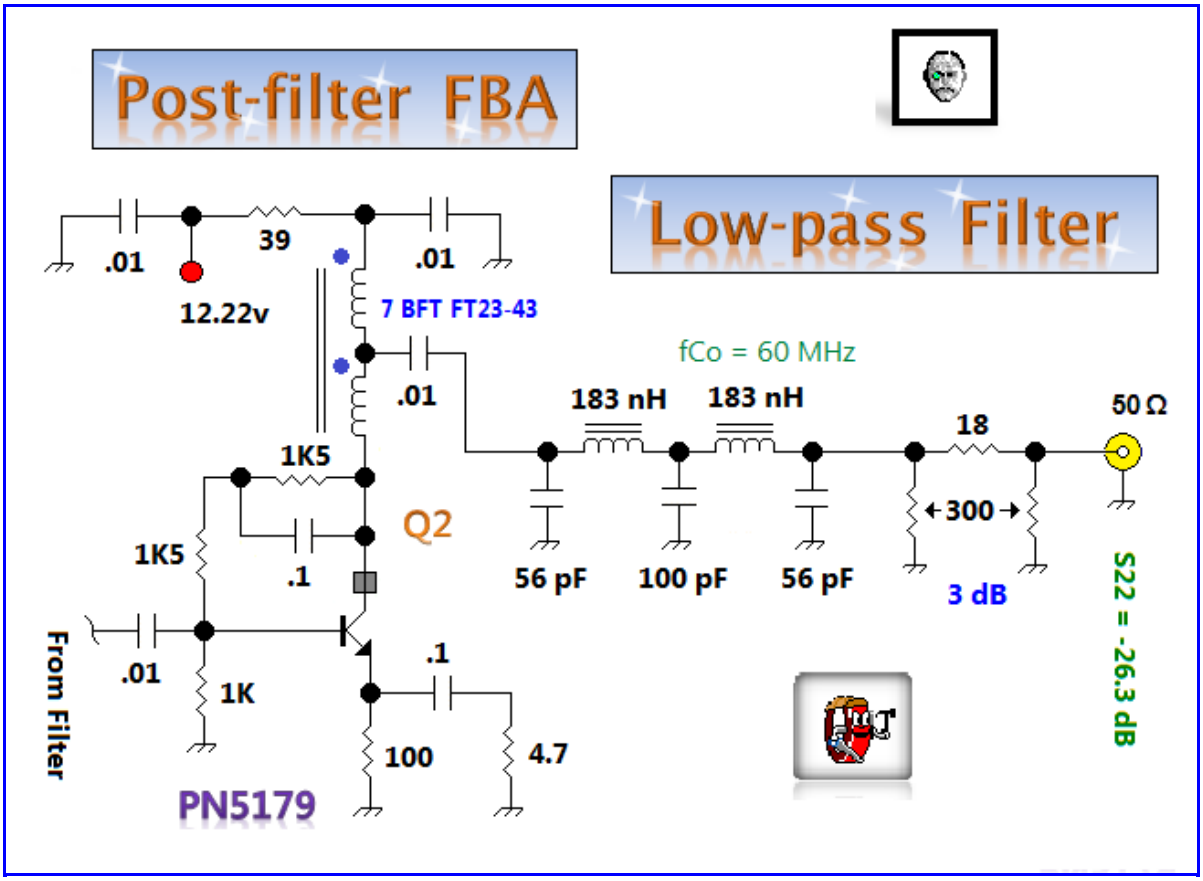
Above — Schematic depicting how to tune up Cy.

Tuning Cy matches the band-pass filter output to the Q2 input impedance — it's fascinating to examine the interdependence of these 50 Ω stages. After setting Cy, I connected the 50 MHz signal generator to the Q1 input and a 50 Ω terminated 'scope to the output and re-peaked Cx, C1, C2 and C3 — finally I tweaked Cy 1 last time with the whole stage in a return loss measurement set-up.



Above — A GPLA simulation of the triple-tuned band-pass filter. CF = 50.125 MHz. I substituted 6 pF (the nominal value) for the 2 series end capacitors in my simulation. отлично!

Post-filter Feedback Amplifer, Low-pass Filter and Pad



Above — Q2, the post filter feedback amp (FBA), an N=5 Chebyshev low-pass filter plus a 3 dB pad.

For Q2, I copied Q1 to deliver a strong input and output return loss. In many circuits employing cascaded FBAs, you increase emitter current in each successive FBA to reduce distortion, however, increasing emitter current affects both the input and output impedance and may trash your amplifier's S11 and S22.

I spent days studying, simulating + bench testing different amplifier designs in the Q2 slot — I generated enough material for another web page and plan to show this work in an update to my Popcorn superhet receiver some day.

It's possible to overdrive Q2 depending on your amplifer power and stage matching. If so, you might consider placing a 3-4 dB pad after the band-pass filter. Some might opt for a 7 element low-pass filter; experiment — as always.

Low-pass filter inductors = turns on T30-10 toroids, although #6 material toroids, or air coils will work fine.

Outputs

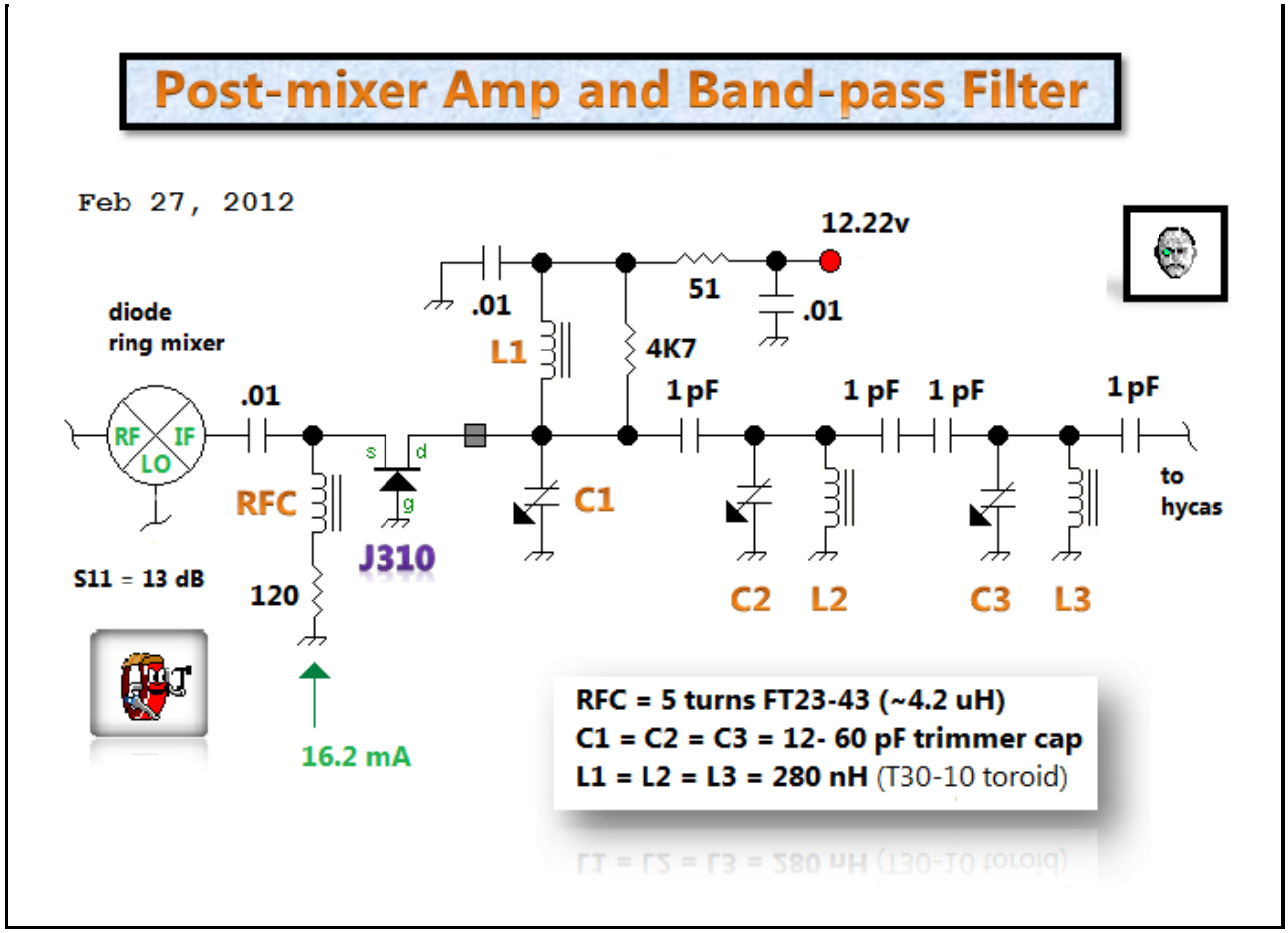
After bolting down the boards, wiring the DC and RF and confirming it worked, I finalized alignment. Using a frequency counter, I tuned the VCO to 50.125 MHz (the half-way point) and peaked C1, C2, C3 for the maximum peak-peak voltage into my 50 Ω terminated 'scope.

Click for the output at 50.125 MHz — 10.09 dBm. I normally hang an outboard 50 Ω attenuator on the output of my signal generators and keep 3 dB, 6 dB, 10 dB and 20 dB BNC-connected pads handy. With a 3 dB pad, the output power = 6.84 dBm — perfect for switching Level 7 diode ring mixers. Click for the 'scope shot with a 3 dB pad applied.

Click or click for an FFT of the output signal. The second harmonic is > 50 dB down. What fun!

The "vestigial" RF gain control shown on the chassis remains unused; wastage.

Miscellaneous Photos and Figures

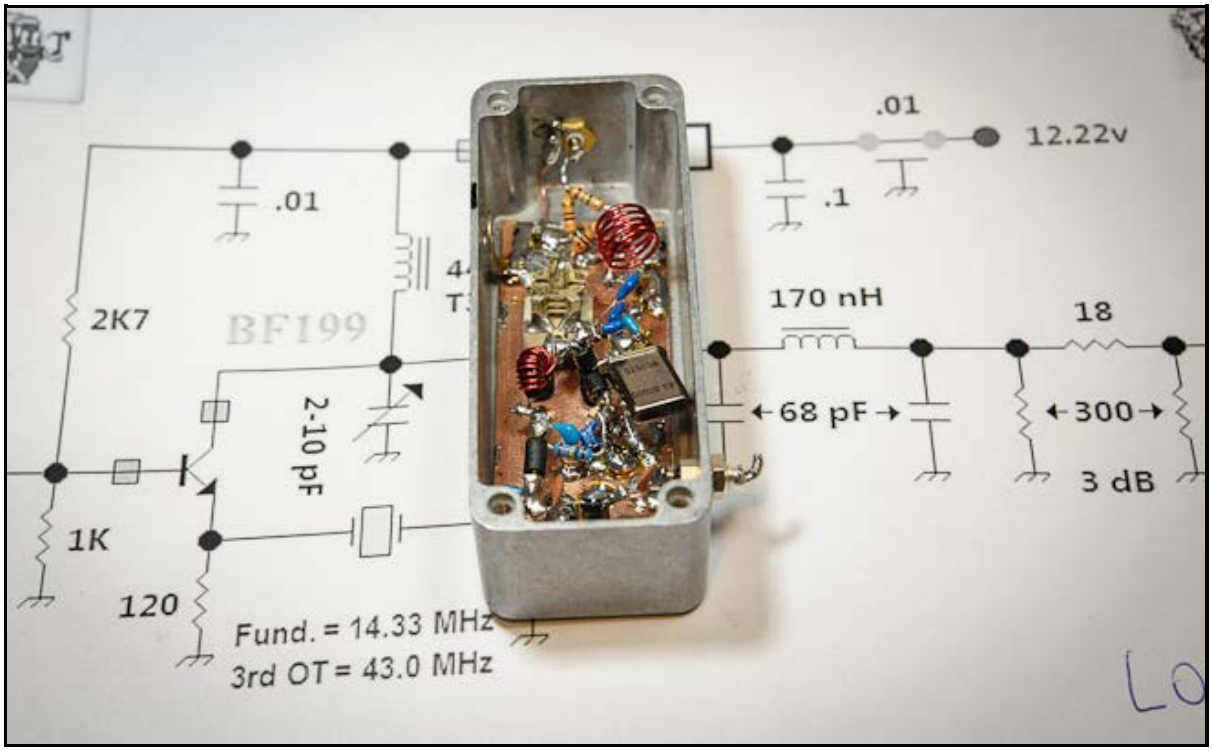


Above — A **failed** experimental JFET post-mixer amp with tuned output driving a double-tuned filter.

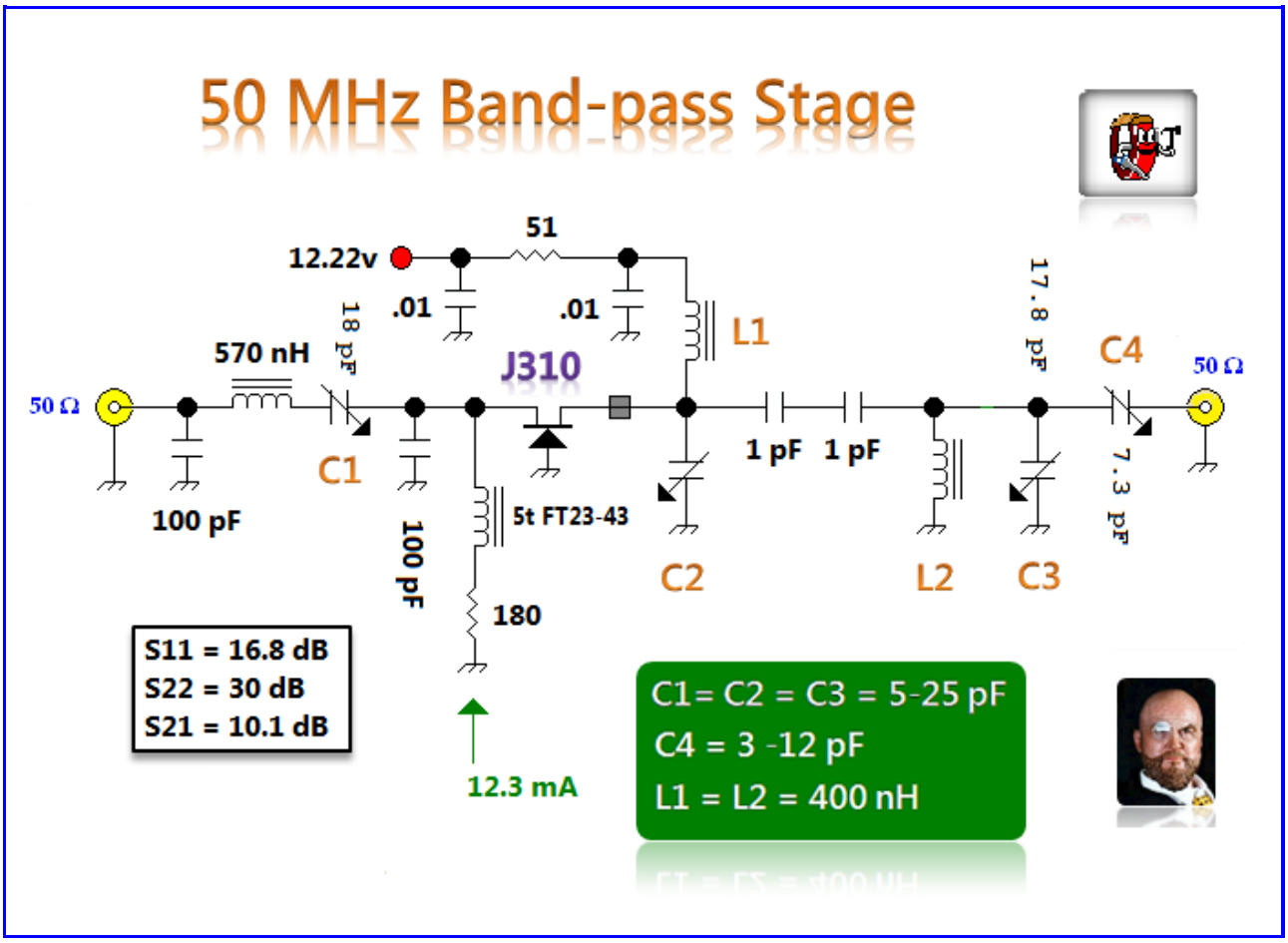
Click for a GPLA simulation of the double-tuned filter. The common gate JFET amp provides a great way to terminate a diode ring and obviates the need for a diplexer network. Click for a breadboard photo of the above stage.

The amplifier input match @ 50 MHz is only ~ 13 dB, however, we're not interested in a narrow band match — the tuned output network makes strong input matching at 50 MHz impossible *(for me at least)* without additional L and C (narrow band components that we don't want!). I tried a few tapped inductor schemes, however, at VHF, adding turns added significant capacitance and things got ugly fast.

The common gate JFET amp/filter goof-up shattered my expectations. The 4K7 input/output impedance drove instability through unwanted coupling between the inductors. I learned my lesson: at or above 50 MHz, *stick to 50 Ω stages* for stability.



Section 4. 50 MHz Receiver Pre-amp and Filter

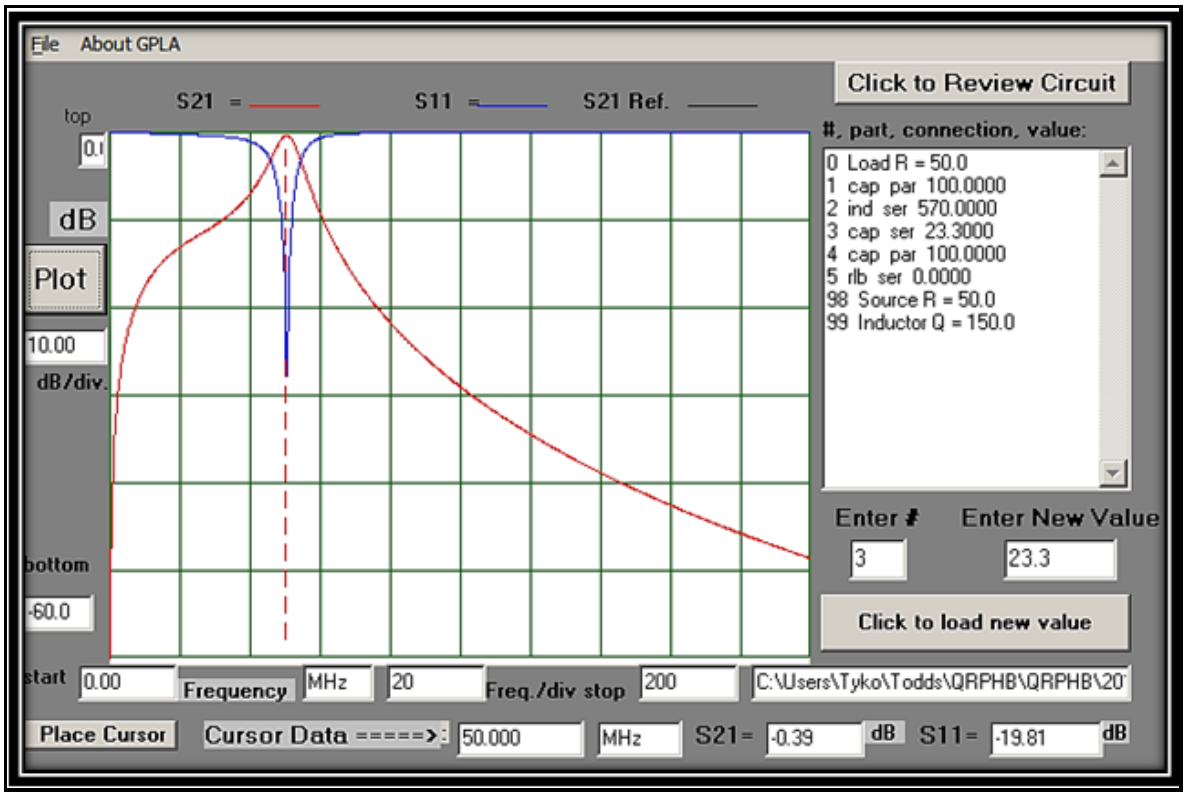


Above — A 50 MHz receiver front end filter with embedded common gate amplifer.

Inspired by the General Purpose Monoband Receiver Front End from Figure 6.69 in EMRFD, I applied inductive and capacitive reactance modeling, DTC08, Ladbuild08 and GPLA08 from the EMRFD ladpac series and built a 50 MHz equivalent.

Connect an antenna to the input and a 50 Ω impedance mixer to the output.

I tested the stage at 50.0 MHz and wound my inductors on T30-10 toroids, although #6 material cores would work okay. You'll find all the measurement techniques in my RF Workbench series 1-4 available though the top-level menu.



Above — GPLA simulation of the peaked low-pass filter "built" in Ladbuild08.

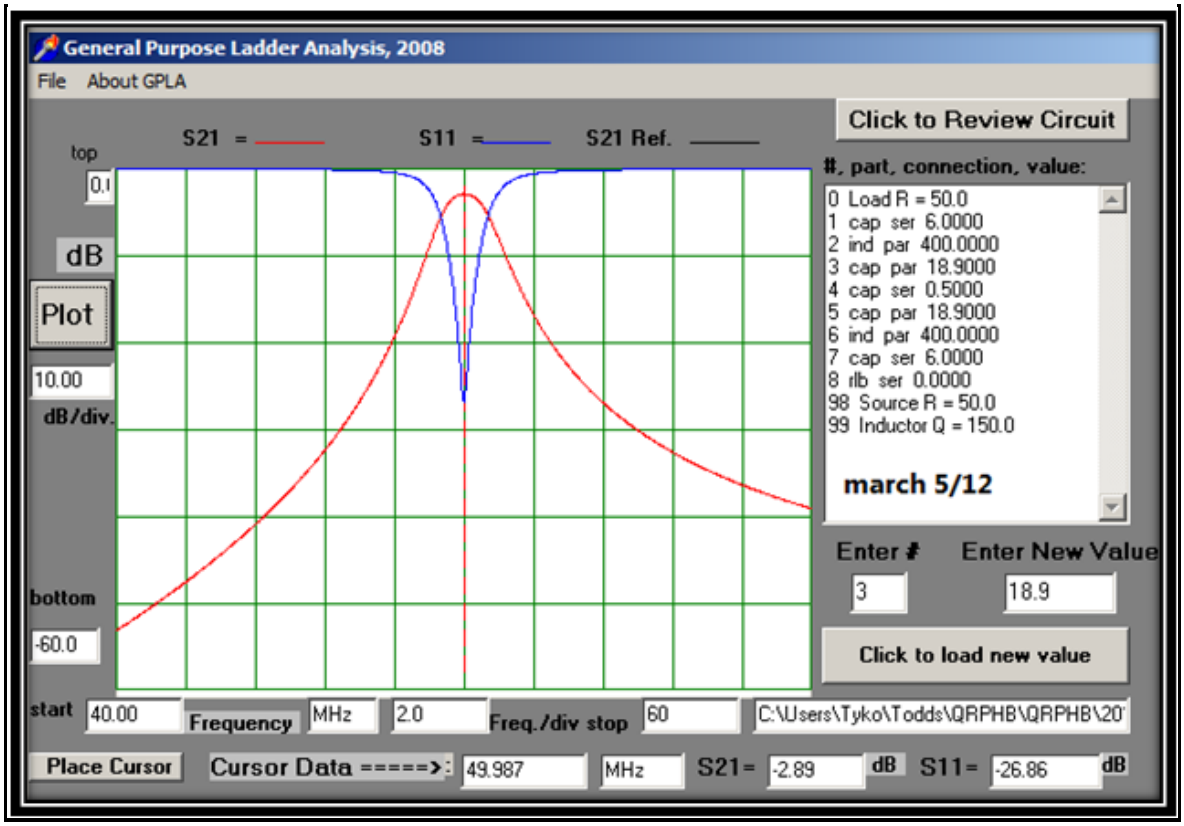
Wes often employs a peaked low-pass filter and after studying his work, I can see why — way better attenuation than a simple 3 element low-pass filter. The FM broadcast band runs from about 87.5 to 108 MHz and in Russia, they call it "YKB" (Ультракороткие волны) or ultra-shortwave. At 87.5 MHz, attenuation = 25 dB; pretty good for such a simple filter. At 144 MHz, filter attenuation rises to ~ 40 dB.

This peaked low-pass filter acts as a preselector for the JFET amp that follows it. *Please read the text describing Figure 6.69 in EMRFD for some great notes by Wes.*

In the simulation above, a 50.0 MHz peak response occurred with C1 at 23.3 pF, while in my real circuit, the capacitor was set to ~ 18 pF. Stray L and C + the input Z of the JFET amp caused this variance, but assuredly; GPLA gets you close.

To peak the low-pass filter, I connected a return loss bridge to the input port and tweaked C1 for the lowest possible peak-peak voltage (tuned for the the best return loss which = 16.8 dB in my circuit). *You may also compress or expand the 540 nH inductor to aid tweaking.*

Since common gate amplifiers often exhibit a lower noise figure with a slight mismatch, an S11 of -16.8 dB works fine .I wish I had the gear to set the input match for the lowest possible noise figure — perhaps 1 day I will.



Above — A GPLA simulation of the 50 MHz double tuned band-pass filter "built" in DTC08. The bandwidth = ~1.8 MHz and varies slightly with the tuning of C4.

I peaked both C2 and C3 with a 50 Ω signal generator and a 50 Ω terminated scope connected to the input and output ports respectively.

Next, I connected my return loss bridge to the output and tweaked C3 and C4 for the lowest peak-peak voltage — the best return loss — and since you tweak 2 capacitors, a strong output return loss delights you.

Finally, I measured the peak-peak voltage with the amp in-line, and after removed the amp and reconnected the 50 Ω cables with a through-connector. Inputting the 2 pk-pk voltages into Applet H on the Design Center web page gave a gain *or S21* of 10.1 dB. I repeated all of the steps above a couple more times to ensure I had set C2, C3 and C4 perfectly.

I found tuning the resonators difficult due to the sharp tuning and wide capacitance range of C1-C3. Assuming your tanks are peaked, the best amplifier gain correlated to the highest input and output port return loss. Have I stressed the importance of a return loss bridge enough?

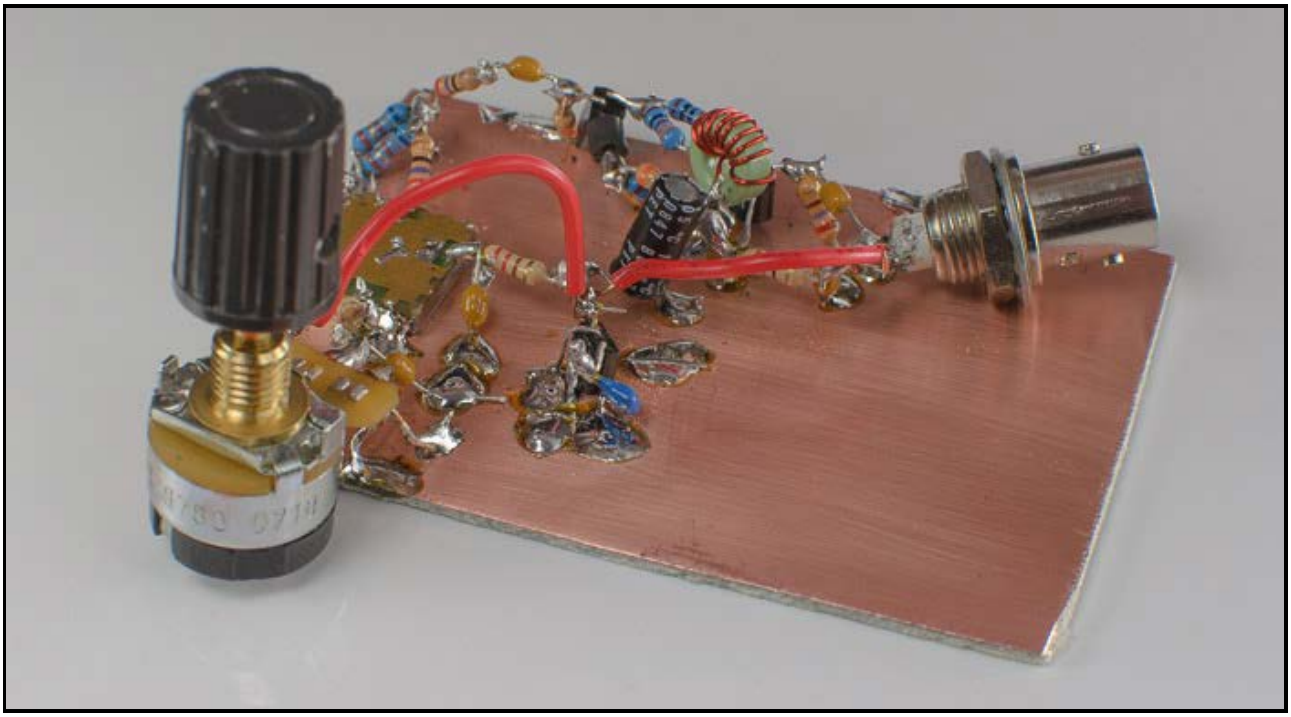
10.1 dB gain should be enough gain for listening to terrestrial 6 Meter band signals with my 5 element Yagi antenna.

Above — A photo of my protoype 50 MHz pre-amp breadboard. In my "keeper" version, I'll swap in a U310 JFET and bias it for ~15 mA.

Section 5. QRP — POSDATA: Z-Communications VCO Experiment

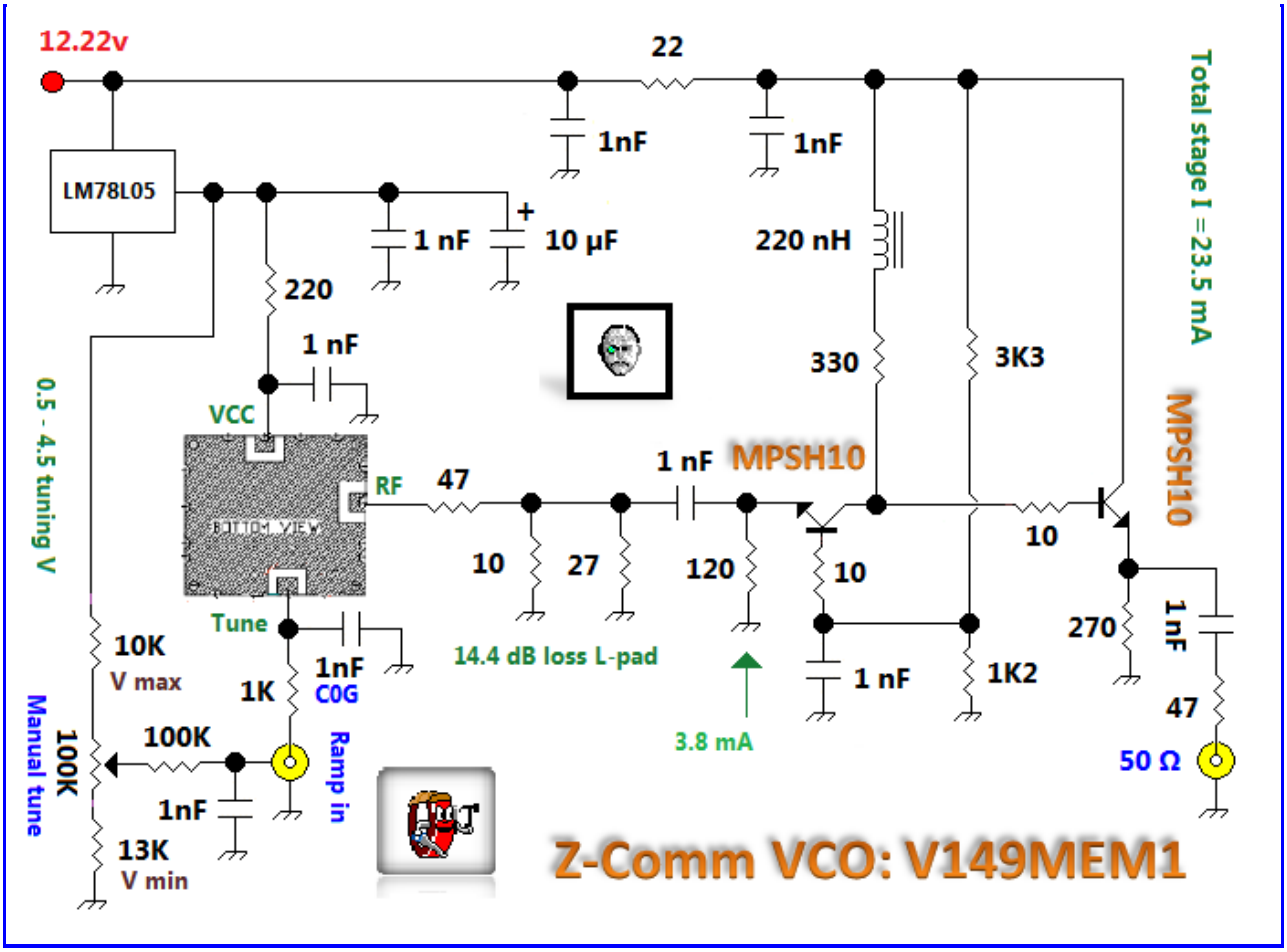
Looking on eBay, sellers list numerous VCOs, although most are surface mount and go well above VHF. My favorite VCO comes from Mini-Circuits Labs: the POS series. Click for an example: the POS-75. These "plug- in" VCOs come in same package as the SBL-1 mixer and are likely obsolete, but still for sale. If you're building a frequency synthesizer with low phase noise requirements, MCL VCOs seem hard to beat. You can still order them from MCL, but the high product and shipping costs might alarm you.

I've looked for cheaper alternatives and the Z-Comm VCO raises 1 possibility. Last year, I purchased a V149MEM1 device for 5 dollars including shipping. Some experiments follow:



Above — My first breadboard. Lacking the MINI-16 receptable, like with MCL POS VCOs: I turned it upside down and soldered the metal case to my ground plane. If I were to keep this circuit. I wound solder all 4 sides to the copper clad board, plus run some copper de-soldering brade from the bottom to the ground plane, or even cut a square hole and flush mounted the VCO on is back.

While mounting it upside down deviates from the recommendations found on the Z-Comm mounting datasheet, I fiigured that for VHF at least, it might work okay. We desire low inductance grounding, but creativity might allow dead bug construction techniques to work.



Above — My complete VCO. The Z-comm VCOs require at least a 10 dB pad on the output to keep port return loss high. Without a pad, you might see something like this plus boost the phase noise. In my circuit, I applied a resistor L-network with ~ 14.3 dB loss to pad the output and provide a match into a common base amp with an input impedance of ~ 6.8 Ω.

The 2-stage buffer is the brainchild of Bob, K3NHI and I love it. This buffer features a common base stage driving a emitter follower yielding high bandwidth and great reverse isolation. Normally, at VHF, the buffer is followed by more such stage(s), or a MMIC. The 220 nH inductor wound on a T30-12 toroid improves the high frequency response of the common base amplifer — experiment with this L to suit whatever VCO you wish to buffer. The gain of the 2 amp buffer is typically around 9 dB and the return loss at the input and output ports lies under 11 dB, so apply attenuator pads to boost S11/S22 as required.

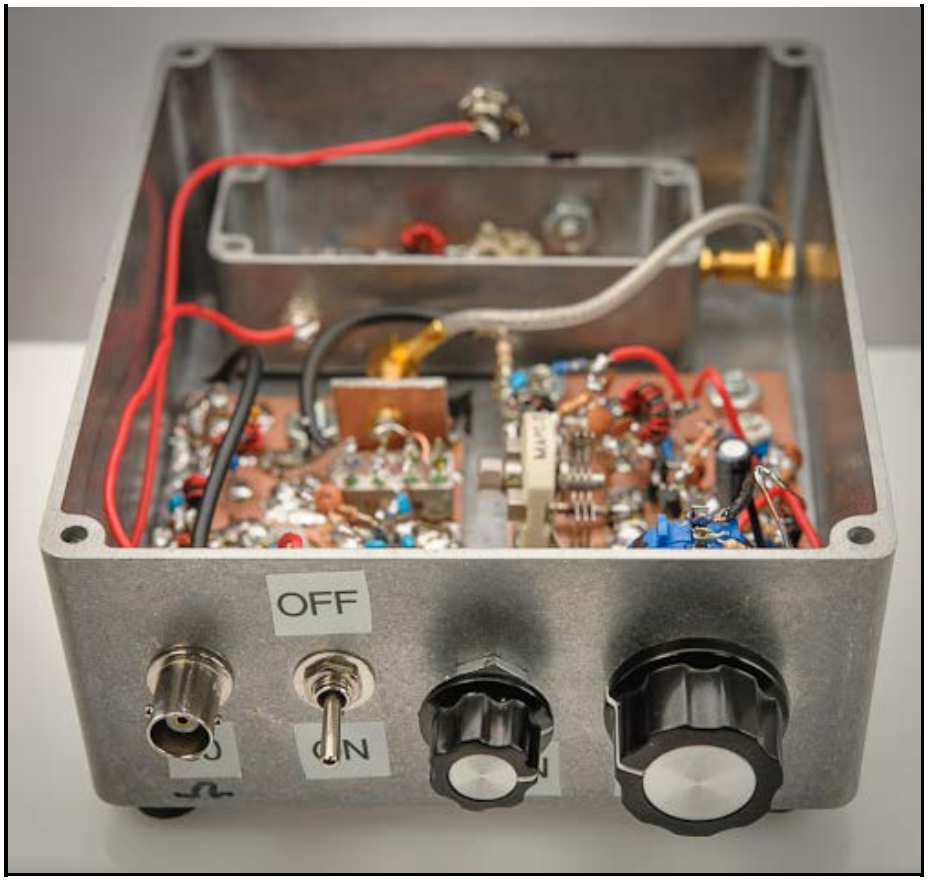
Click for the scope tracing at 0.5 VDC tuning voltage. Click for the 4.5 VDC tuning voltage 'scope tracing. The harmonic distortion at the lowest tuning voltage = ~ -19 dBc and decreases to -28 dBc at the highest tuning voltage; better than specified. Notice that power decreases as frequency increases. All the commercial VCOs I tested do this. A higher fT amp like the PN5179 or other BJT might be a better choice to offset the power change versus frequency *contribution* of the buffer/amp.

For a sweep circuit, I would mix this VCO with another low level, single frequency VCO with its current controlled by a downstream leveling circut to derive a flat amplitude over the range of the VCO. I plan to try the Z- Comm V150S015 in such an arrangement to make a 70 - 150 MHz VCO for sweeping.

Please refer to the datasheet for the pin out on the Z-Comm VCO: I chose the pinout shown in the schematic to make an efficient drawing. The two 10 Ω resistors in the buffer/amp snub UHF oscillations first measured by Bob and confirmed by me. Ferrite beads might work as alternates.

Section 6. Miscellaneous Photos

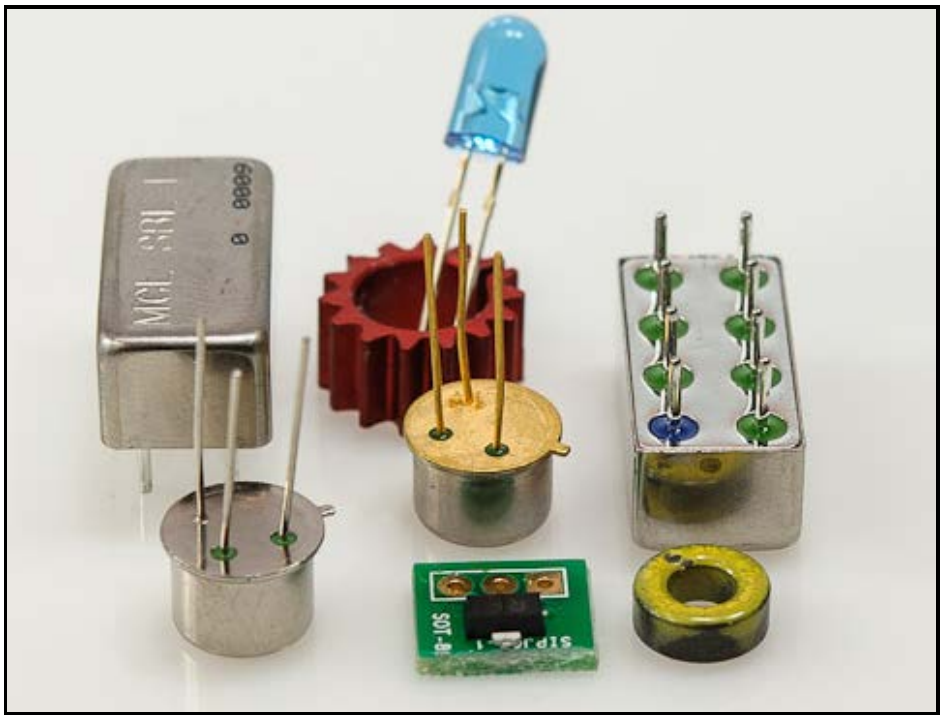
VHF SIGNAL GENERATOR
138-172 MHz
98-109 MHz
50Ω
ON
OFF
ON
50Ω


OFF
ON

TERMINATOR

RF — Test and Measurement

HF Ragbag



It's easier to present short topics on catch-all web pages — HF Ragbag shows some 2012 non-VHF experiments in no particular order. I also share thoughts on circuit building and writing: we can think and work better.

1. Comments from the Workbench - The Need for Clarity

On Building

In 2012, I boosted my circuit and writing quality to improve your experence: a genuine, return-to-basics approach in amateur, component-level electronic design. As possible, RF circuits will feature 50 Ω input and output ports — totally adopting a 50 Ω environment — for I'm convinced this is the best way to go. The 50 Ω building and measuring standard offers much: an easy-to-interface modular approach; 10 dB improved sensitivity over a 10X 'scope probe and if wanted, measurement with commercial or homebrew test equipment such as a spectrum analyzer, network analyzer or RF power meter.

Like many, I started out by collecting and copying circuits with little emphasis on true understanding. I wanted a completed circuit — quickly as possible — failing to develop my design skills. Without design skills honed by studying and properly measuring our circuits, we bide in hit-and-miss electronics — a frustrating repetition of trial and error, over and over again. We ought to adopt the attitude and thinking of engineers while keeping our design work —including the math — fun. However, embracing scratch-homebrew electronics with the overall goal of trying to understand each stage takes effort. "There is no substitute for hard work" wrote Thomas A. Edison. Scratch homebrew involves reading, simulating, collecting parts, mastering new techniques and building or buying test equipment.

This is more than knack, an abused noun that often means "hack". Our key tasks: to measure, analyze and understand the circuits we copy or create takes patience and practice. Dissecting circuits to understand their function means to hypothesize and reflect — to apply science on

paper, with software, and finally, through careful bench experiments. Often we lack the math skills or test equipment to fully investigate some aspects of our circuits, but try our best: measure what we can measure, seek help and grow. I hope this site shows our hobby can be less about making stuff and more about the rewards of actual design: an authentic, personal journey to get better at something you love.

I've never been much of a kit builder; it's too much like Max Klein's Paint by Number for my tastes. But kits dominate HF QRP homebrew and may offer a cost effective way to make gear; especially test gear. Stuffing parts in a printed circuit board won't teach you much about design, but might get your feet wet. Some people remain perfectly happy building kits or madly copying circuits — all the power to you! Do whatever you want. One day you might awaken, but don't worry; I won't try to goad, or convince you.

My favorite builders include people over 50 who suffer the often crippling symptoms of 'appliance apathy' — an epiphany reminds them why they first got into radio: homebrew experiences. Maybe a crystal radio set, or a simple superhet receiver they breadboarded long ago. *Then they come back full circle*; like a loop antenna. Oh-boy — "Bob" rediscovered his radio roots and needs to unleash his creativity and passion to learn and improve. I write for people like Bob. Heck; I am Bob.

On Writing

You'll notice improved narrative writing too: I prefer to read and write crisp statements in short sentences and paragraphs. Brief, yet descriptive text accompanied by ample white space, clear headings and bulleted lists invites you to read on. Plain language writing — simple, clear, writing that is easy to read and understand — signals a refreshing move away from the turgid, word-filled claptrap I learned in grade school. Making your prose easier to read requires greater effort writing and re-writing. My first..to...fourth drafts always suck.

Passive verbs, or nouns and adjectives that function as verbs with no clear subject confuses readers and boosts wordiness: I employ active verbs to invigorate my writing — active verbs connote me or some else performing an activity you can visualize or feel. Actions that may inspire, persuade, or even vex you! Ours' is an emotional hobby.

RF electronics contains rich amounts of jargon. Of course, we must learn some jargon to communicate our ideas as hobbyists, but writing jargon to impress, or to place yourself above others lacks humility and alienates people. Do you know anyone who likes being talked down to? The first step towards becoming humble is to admit you're not humble and then work on it — and I'm working on it.

Although I enjoy writing about electronic experiments, I'm not sure it's worthwhile — Does anyone actually design circuits anymore? Well, back to my 1970's-style analog experiments...

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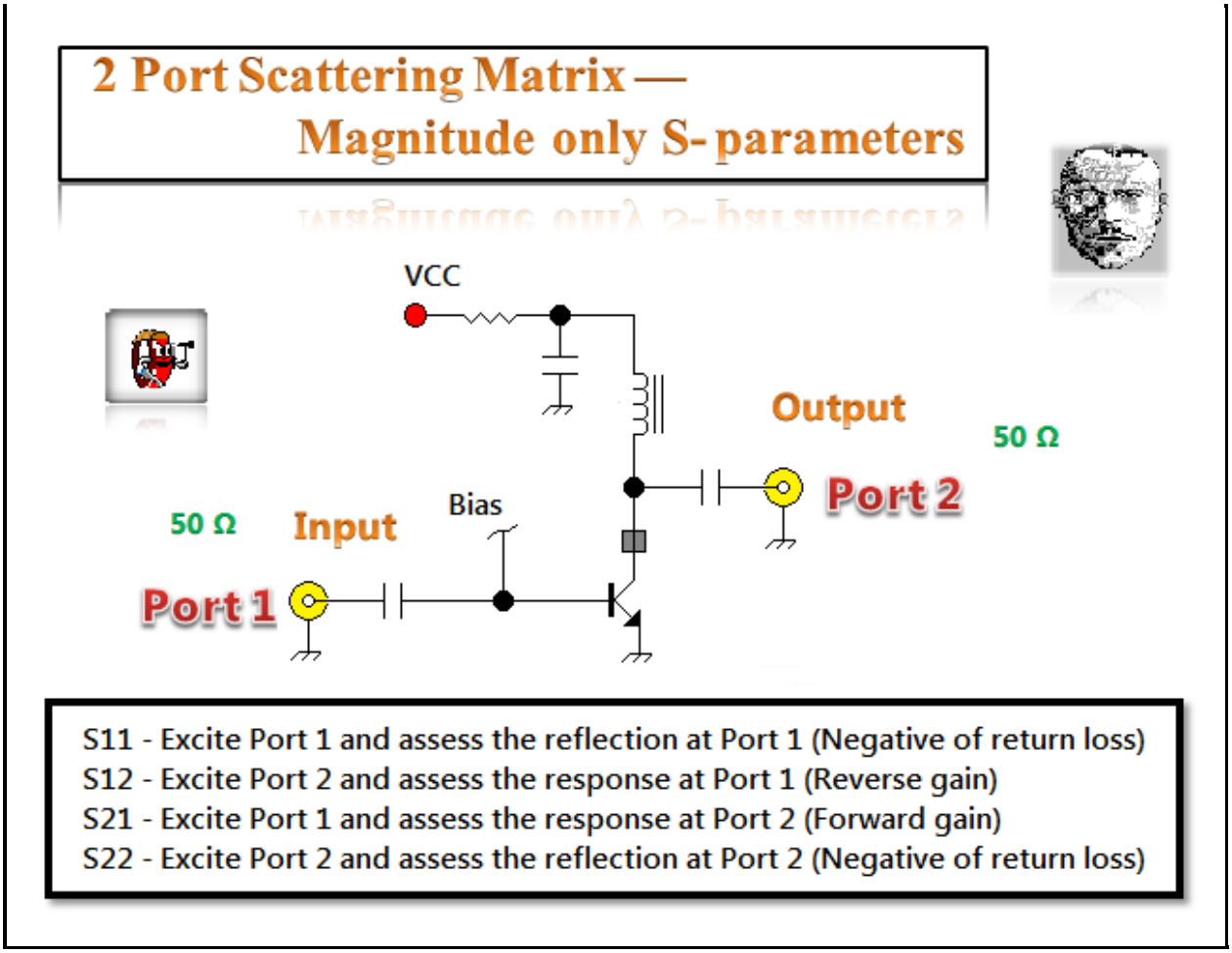
2. Magnitude Only Scattering-Parameters



Above — A simple model describing the S-parameters displayed on QRP / SWL HomeBuilder in a Class A amplifer with 50 Ω ports.

S-Parameters

Any device with 2 connectors may be modelled at AC for a specified frequency with just 4 scattering parameters: forward gain, reverse gain + input and output impedance (match or VSWR).

S-parameters address voltage ratios: comparing the amplitude of different signals at the 2 ports. For example, S21 is the magnitude of forward gain and equals the ratio of output voltage to input voltage.

S-parameters are vectors; a mathematical quantity that may be visualized as an arrow anchored at 1 end that pivots around its base. The length of the arrow represents magnitude while the angle it makes with another vector or its base line decribes its phase in degrees. In addition to phase and magnitude, S-parameters allow analysis of gain, stability, complex impedance (resistance + reactance), admittance and other vector quantities.

Measure S-parameters with all ports terminated in a 50 Ω impedance.

Some of us worry only about the gain, losses or "match" in our 50 Ω circuits and could care less how the signal phase changes as it passes through our amplifiers or attenuators. I express only S-parameter magnitude in logarithmic form (dB) and take this *Über simplified* approach because builders can easily measure S11, S12, S21 and S22 on a 50 Ω test bench with a small staple of bench accessories + a 50 Ω 'scope or detector.

Topics like matrix theory, vector math, the "jay" operator, converting S-parameters into other matrices, Smith charts etc. may turn *off* the average amateur designer. You advanced readers, may raise your 2 port network skills by visiting better web sites + reading books, simulating with SPICE, or better yet, measuring your port parameters with a vector network analyzer.

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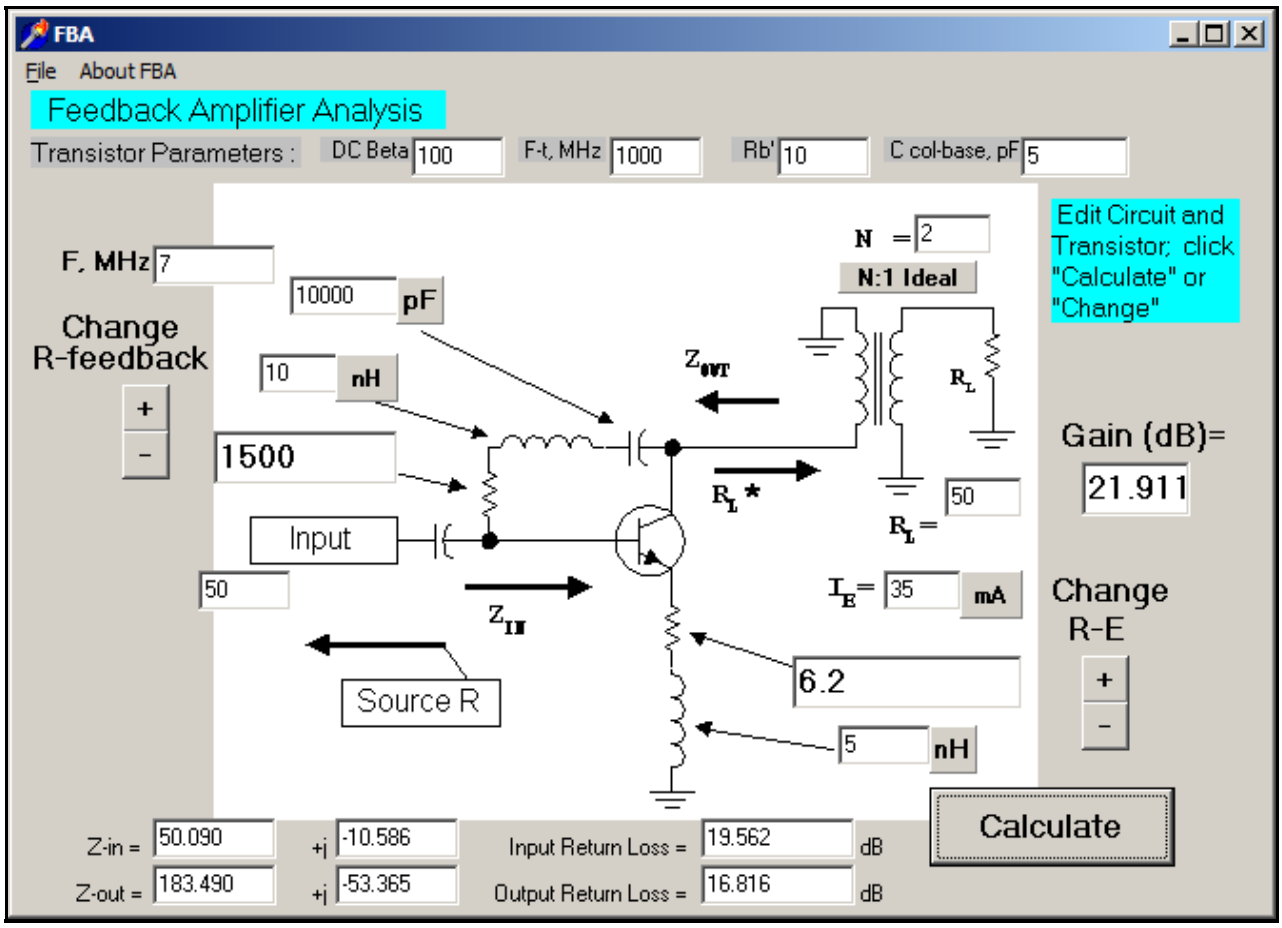
3. More on Feedback Amplifiers (FBA)

Many builders (myself included) copy feedback amps rather than design their own. By tweaking the emitter current, shunt and series feedback while measuring S11 and S22, plus simulating with a program called FBA08.exe, I've learned it's possible to design good feedback amps FBA08

is 1 of the Ladpac programs that ships with EMRFD.

I wanted a FBA with ~35 mA emitter current for improved IMD and low distortion on strong signals. Such an amp might follow a diode ring mixer in a receiver I.F. chain.



Above — My 7 MHz FBA set up. Wes, W7ZOI suggested using 5 nH as the default emitter inductance and 10 nH for the default collector to base inductance in FBA08. These represent stray inductances in your circuit breadboard. Emitter inductance affects the input impedance more.

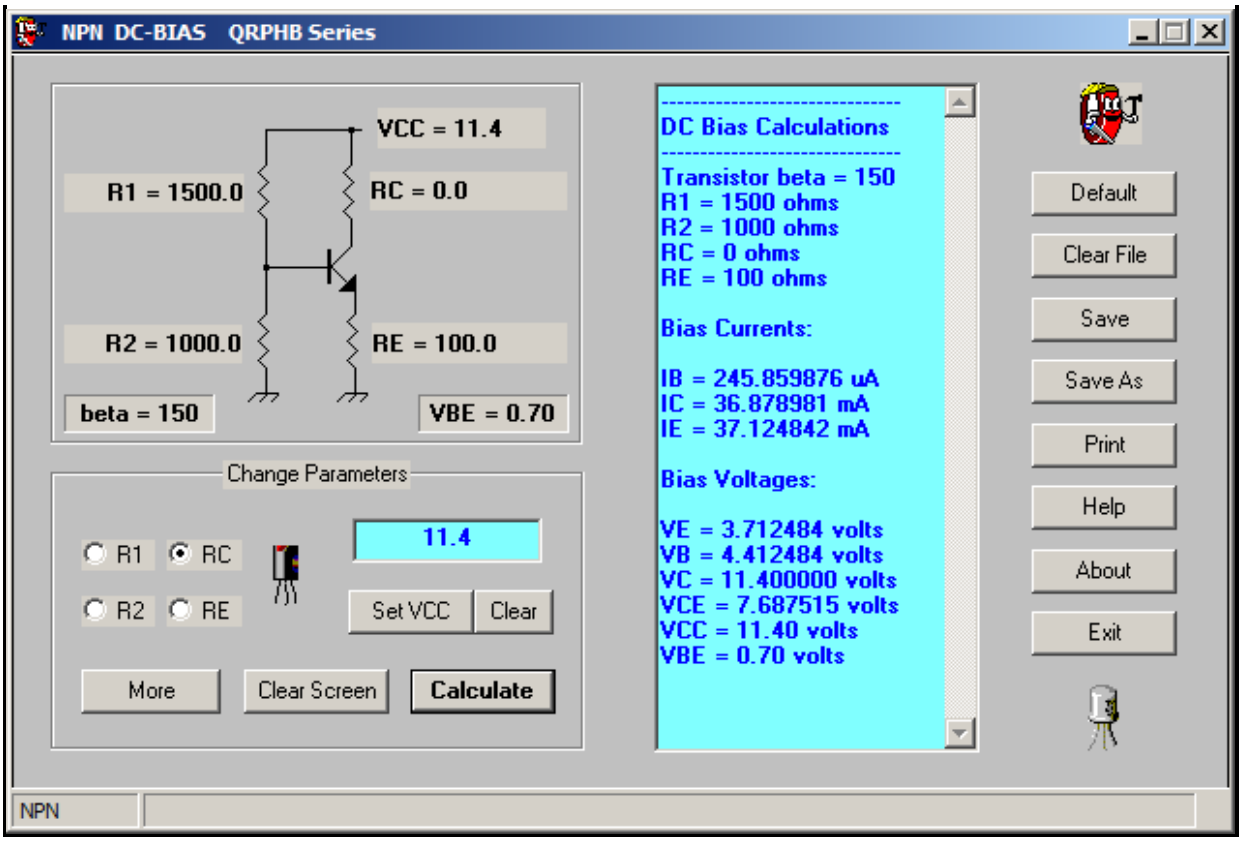
Zin = input impedance. Zout = output impedance.

Explore this program to learn how changing the emitter resistor, feedback resistor and emitter current affect the input and output return loss.

Adjusting the transformer N and load values only affect the calculations for Zin because this app wasn't really designed to crunch output transformer Z ratios for Zout manipulation. The default output Z = 200 Ω and thus for the N parameter with a 50 Ω RL, RL is multiplied by N^2 to set the amplifer load impedance.

From FBA08 simulations: with an emitter current of 35 mA, my series feedback = 6.2 Ω and shunt feedback = 1500 Ω.

I chose a simple voltage divider bias network to set up the ~35 ma and ensure reasonable temperature stability.

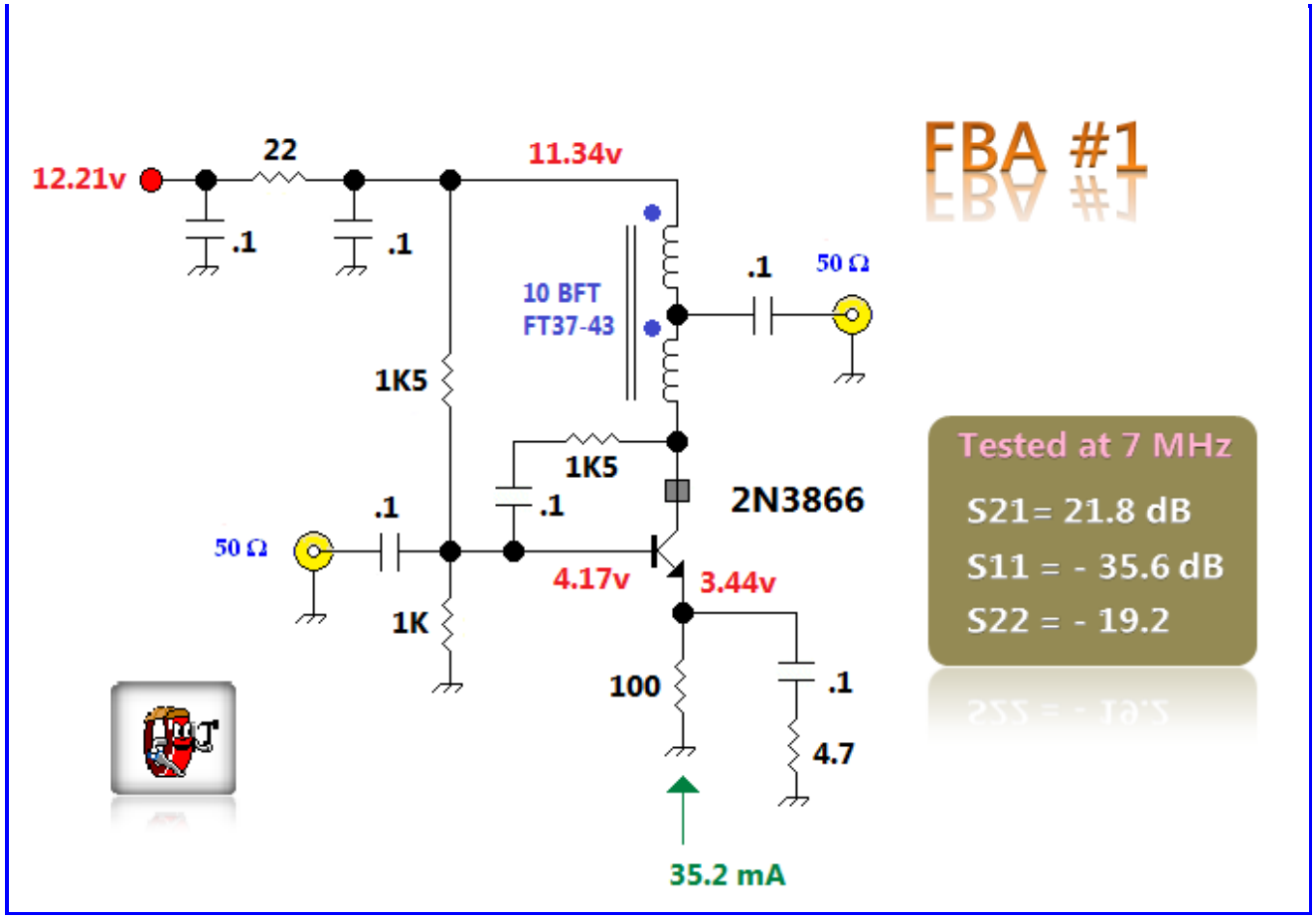


Above — Choosing the emitter and nearest standard value bias network resistors to set up ~ 35 mA emitter current with a program. Actual biasing requires you to set up the correct emitter current + establish reasonable temperature stability.

Click and scroll to #5 for some basic transistor biasing notes. While this supplement shows a simple method for stable bias networks, it probably understates that **Beta** bias stabiility is a function of the ratio of RB to RE, where RB = the 2 base resistors in parallel. The lower the ratio the better, but then more input power is lost in those resistors. A higher ratio reduces stability but wastes less input power — another trade off we must negotiate! See Ken Kuhn's web site for thorough, expert-level information on voltage divider biasing your BJT amplifiers.

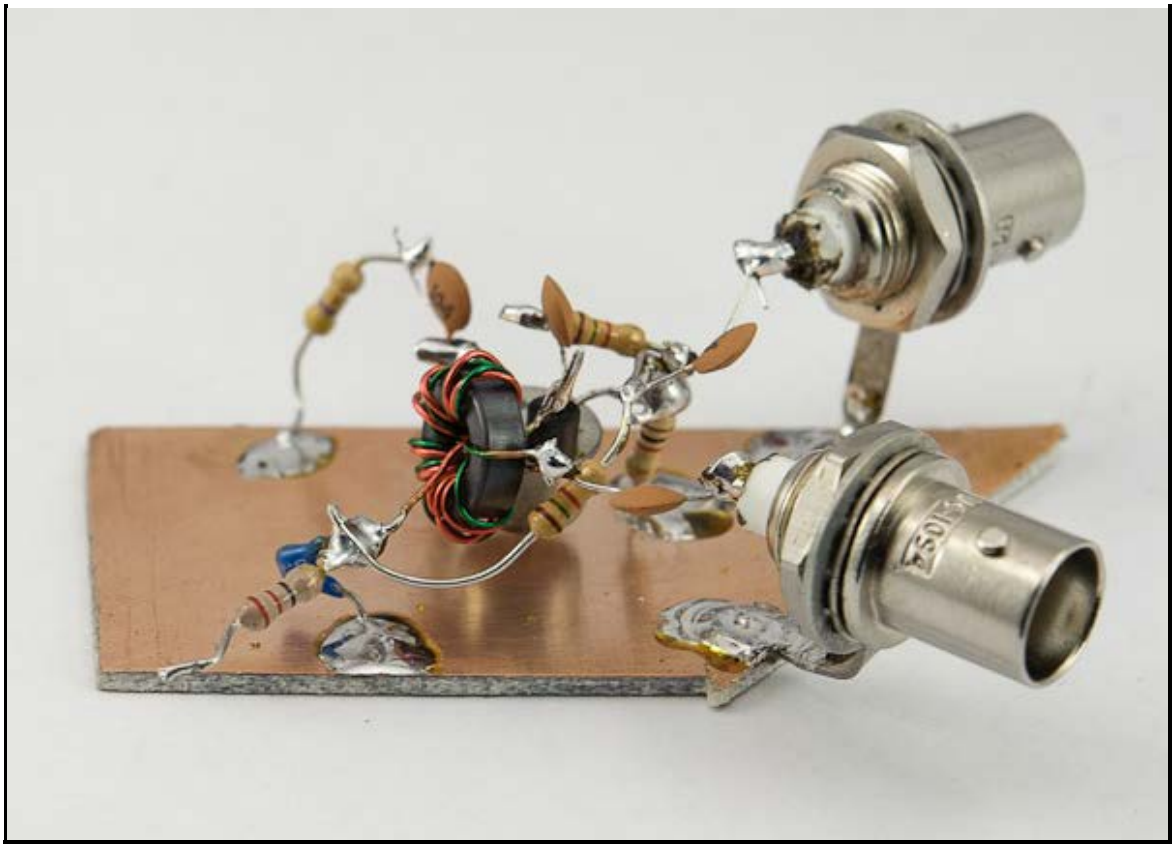
I use NPN DC BIAS, a program I wrote, however, Wes included 1 in the Ladpac software called Biasnpn08.exe that's also good. Determine the VC for the program by first multiplying the value of your decoupling resistor by the emitter current in Amperes to learn the voltage drop across the R. Then, subtract that voltage drop from your power supply voltage: 12.22V - (.0371 A X 22 ohms) = 11.4 VDC.

Our software allows you to pick approximate base and emitter resistor values to set up a desired current in your amplifier breadboard, but you must still choose reasonable values for temperature stability. Tweak them as needed, or choose some other bias method such as a current source. Let's move to the bench...



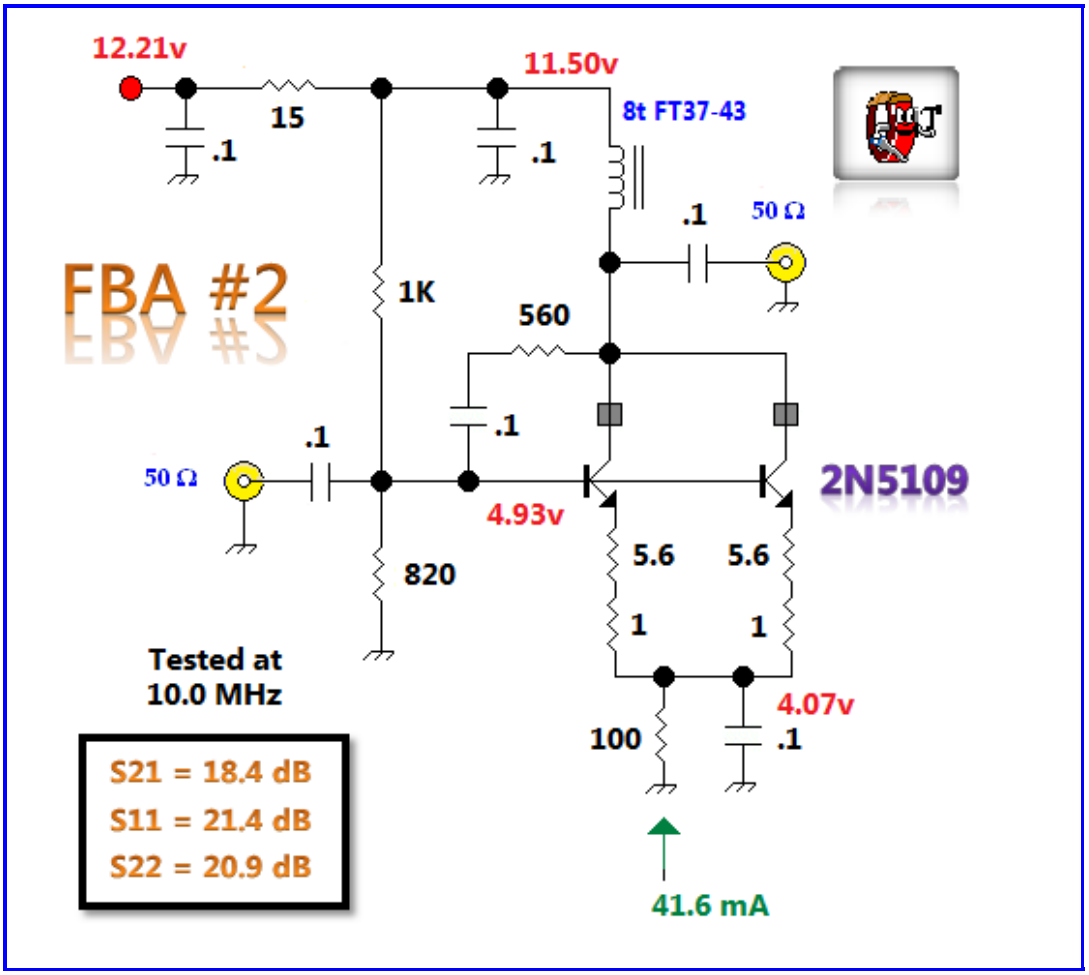
Above — My 7 MHz FBA with some measured S-parameters. On the bench, I lowered the 6.2 Ω series resistor to 4.7 Ω because the voltage divider bias network also affected Z in. I tried 3.3, 4.7 and 5.7 Ω resistors for series feedback and settled on 4.7 Ω since an S11 of -35.6 dB wins the prize!

The S22 of -19.2 dB bettered the value predicted by FBA and seems quite acceptable considering we normally follow a FBA with a 6 dB pad that raises the output return loss another 12 dB. FBA08 gets you close, however, only bench experiments will realize the amplifer you want, and sometimes, a decent S11 and/or S22 may elude you.



Above — A photo of the 35 mA feedback amp built on scrap of copper clad board.

Parallel Transistor Feedback Amp



Above — A feedback amp with two 2N5109 transistors wired in parallel. Click for a photograph of this prototype. I lacked 6.8 Ω resistors and placed 1 Ω + 5.6 Ω to make the needed R for a strong S11.

Originally, I built FBA #2 with a 4:1 Z transmission line transformer, but measurements of S22 disppointed me. Later, a L wound with 8 turns around an FT37-43 ferrite toroid drove an S22 of 24 dB, but S11 was only 18.5 dB.

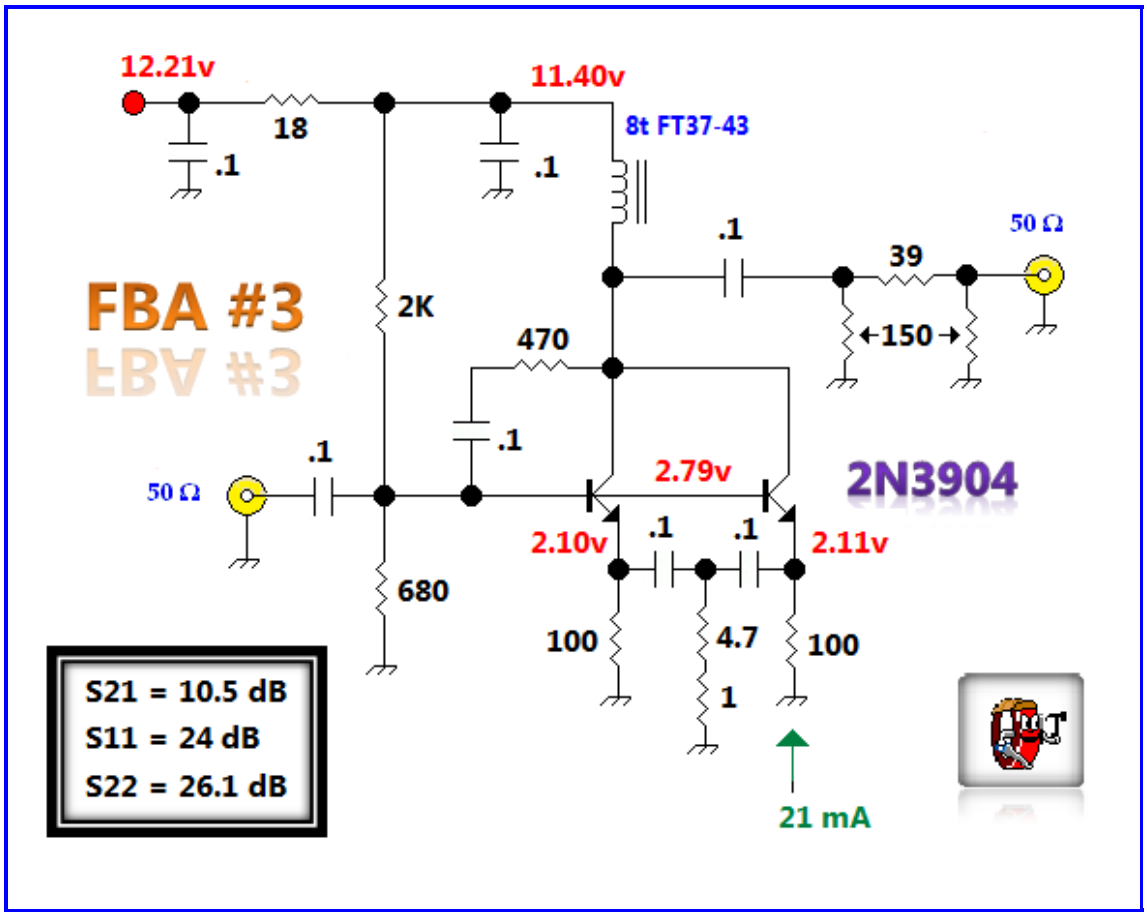
With the amp set up to measure return loss on the input port, I placed a 500 Ω potentiometer in series with a 100 Ω resistor between the collector and base terminals and tweaked the pot to obtain the lowest peak-peak voltage in my 'scope (lowest return loss). After, I removed the pot and measured its resistance at 572 Ω. Finally, I soldered in a 560 Ω resistor and re-checked S11. Perfect. With my goal of at least 20 dB for S11 and S22 obtained, I powered down my bench and took some photos.

Wes, W7ZOI displayed parallel transistor FBAs in EMRFD and other works and recently I noticed Lyle, KK7P employed a parallel NE46134 FBA as a post-mixer amplifer in the Elecraft K3.

Wes wired 2 parallel 2N3904s to avoid using an expensive medium power BJT like the 2N5109. Doing so splits the heat between 2 devices, but does not deliver better IMD performance beyond what is offered by increasing the emitter current. In a typical FBA bias setup, you may measure as much as 10 volts between the collector and emitter terminals and with a supply of 12 VDC + a standing current of 20 mA, the collector dissipation = ~200 mW. This is about maximum for a TO-92 device likes a 2N3904, but only half of maximal dissipation for 2 in parallel.

Then, too, the K3 applies 2 parallel medium power BJTs get power dissipation with an SMT transistor. For strong IMD performance, Lyle and crew are throwing 80 mA or so into the pair — hard to do with SMT parts, so they overcome heat and power dissipation issues with 2 devices. Cool (literally).

Heat sink BJTs when you crank up the emitter current.



Above — An attempted 2N3904-based parallel feedback amp. Each BJT draws ~21 mA emitter current. Without the 6 dB output pad, the output return loss = 14 dB — I *failed* to realize both a strong (raw) S11 and S22.

The power gain including the 6 dB pad = 10.5 dB. I'll discard this design since it's substandard — without failures, victory may taste bland.

For bench designers, making a parallel FBA where both the raw S11 and S21 are > 20 dB is difficult and bench failures may either frusturate you, or enhance your resolve to succeed. With success, great satisfaction arises and I'm addicted to that feeling.

An FBA bench triumph means you managed to establish the perfect combination of series + shunt feedback, emitter current and the correct

output transformer ratio for *that* transistor plus its biasing circuitry — no small task.

A well matched amplifier = a thing of beauty! The fetching trio of high S11, S22 and S21 rewards your efforts and boosts your confidence to experiment further. And so it goes...

Sadly, only a fraction of hobbyists create and evaluate their own circuits.

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4. Microphonics in Direct Conversion Receivers

LO = local oscillator or VFO. DC Receiver = direct conversion receiver.

Microphonics are induced electrical responses that arise from a mechanical vibration on the DC receiver chassis or circuitry. The audio amp, acting like a transducer, makes a clicking, or popping noise when you do things like tap the chassis, or unplug components — the disturbance throws out a burst of DC voltage that's amplifed by the AF chain and pops the speaker.

We may read or hear inexperienced builders tell us to expect microphonics in our DC receivers — *de trop* folklore strikes again! As a student of EMRFD and those wise designers who live in and around Beaverton, Oregon, I share some of their best tips to decrease microphonics in your DC receiver projects. " Keep Your LO From Radiating to the Outside World and Keep Unwanted RF from the Outside World Getting Into Your Receiver" seems the appropos title for the bulleted notes that follow:

Read EMRFD pages 8.7 to 8.11 and then build or apply the presented examples. Wisdom is experiential; it comes by doing, not just reading. It's no accident that Chapter 8 author Rick, KK7B mentions microphonics and hum in the same section. I've never read more thorough notes regarding DC receiver nuances anywhere; for example, did you consider that an ungrounded air variable capacitor shaft poking outside the LO box will radiate LO signal per Figure 8.18 ? I didn't in my early days.

Stick your LO in a RF-tight enclosure with RF-grade connectors and coax to patch the AC signal to the product detector. Bypass RF with feedthrough capacitors on any DC voltage lines that pass through the LO chassis wall. Many enthusiasts have only operated kitted or homebrew DC receivers where the LO and receiver guts lie on the same circuit board — this *ensures* microphonics. Wes and Roger built the historic Ugly Weekender VFO, transmitter and receiver in seperate boxes — resulting in low microphonics and no pulling of the VFO when keying the transmitter. Nothing in that 2 part QST series was done by accident. Read these articles to "go to school".

Reciprocally important; keep unwanted outside world RF from getting inside your DC receiver! Apply resistors plus capacitors, or inductors plus capacitors to decouple and bypass RF from moving along on your DC voltage lines, key line, microphone cables etc.

Keep product detector port-to-port isolation high. Typically, we employ double balanced mixers to obtain high port-to-port isolation. I cover mixer balance on this page . For diode ring mixers, measure the return loss of the circuits that you connect to the product detector LO, RF and AF ports — I aim for 20 dB or greater return loss on my LO output, RF output and AF amp input circuits to help preserve the product detector balance and keep port isolation as high as possible. Along with 50 Ω amplifers, attach attenuator pads, AF diplexers, or whatever to help increase port return loss as required.

LO-RF port isolation: Consider a common gate amp with an output matching network to get a high output return loss (S22). The common gate amp provides strong reverse isolation without adding much noise.

Avoid end-fed wire antennas where there is a strong antenna field right next to your radio.

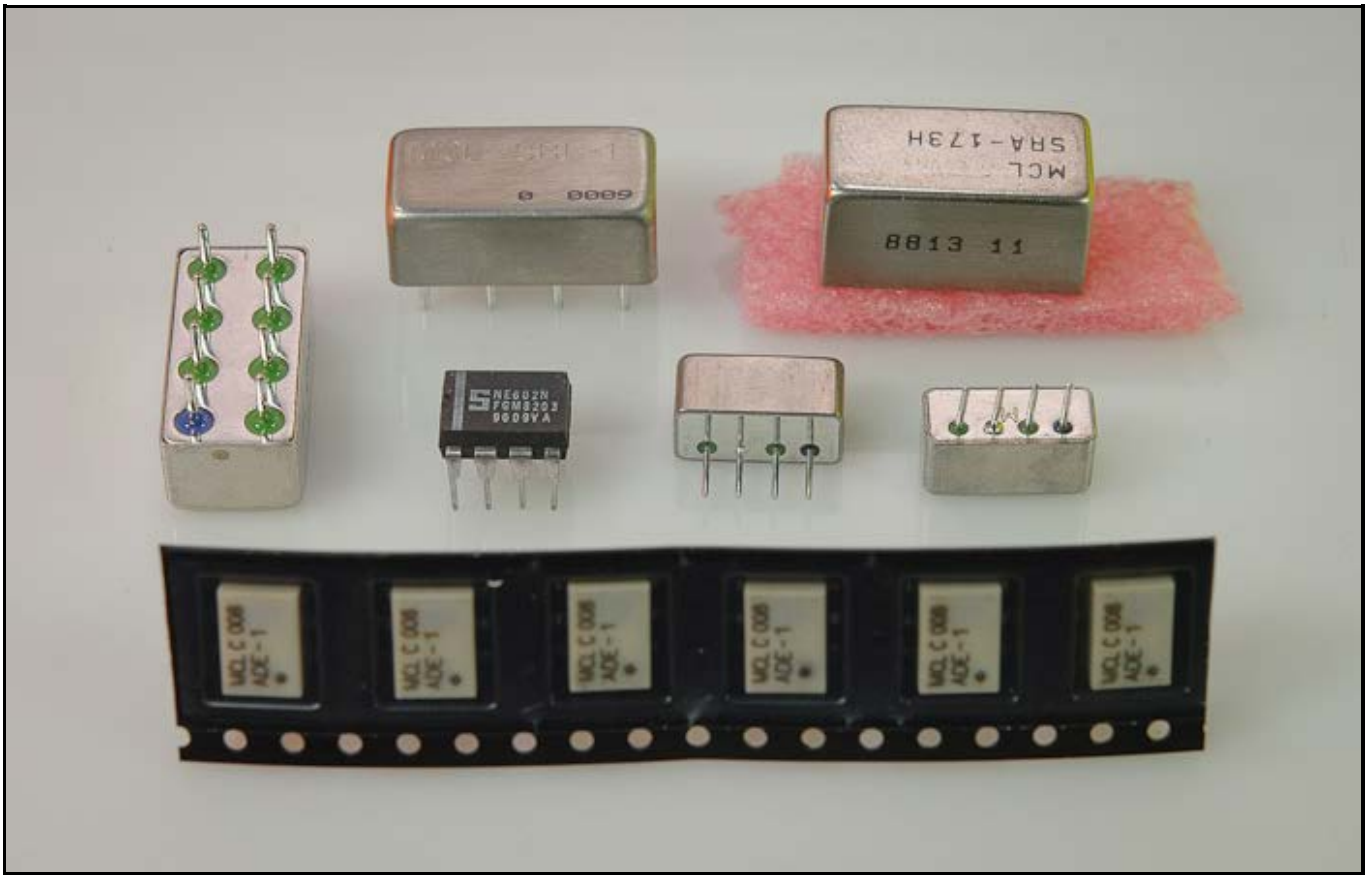
I favor sturdy chassis/cabinets with rubber feet. Homebrew copper clad board or die-cast aluminum cabinets may work best as joints and screwed connections won't corrode. *This is a weak recommendation.*

Double the LO frequency or apply a heterodyne VFO. Often microphonics arise in the VFO tank. EMRFD cover this well. If the VFO operates at a significantly different frequency than any of the signals reaching the balanced mixer, leaked LO won't cause as much havoc as when a LO tank is tuned to the mixer RF port frequency.

Despite proper techniques, RF can exit via the antenna port and make its back into our rig through power supply cables (often modulated by our house AC electricity). In some cases, we require special power supply decoupling to decrease hum and microphonics. We might need to add a common mode choke (+/- capacitors) for common mode noise suppression in addition to the usual differential mode choke(s) and capacitors. In my main shack power supply, I run a common-mode choke plus I soldered a 0.01 uF capacitor across each bridge rectifier diode to bypass RF.

Some radio operators just run battery power supplies.

Above — Feedthrough capacitors. I prefer hole mount over solder mount parts, however, quality feedthrough capacitors of any kind tend to be expensive. As a hobbyist, I'm constantly searching for bargains and when I find 1, I'll purchase a bunch to meet my current and future needs.



Above — Some double balanced mixers from my collection: ADE-1, NE602, TUF-1, TUF-2, SBL-1 and a SRA-173H; a MiniCircuit Lab's Level 17 diode ring mixer.

You owe it to yourself to listen to a DC receiver designed and built to reduce microphonics — music to our ears.

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5. Some Experiments with RF Bypass Capacitors

Introduction

Bypass implies a low impedance path to ground for RF at 1 or more frequencies. After reading EMRFD pages 2.28 - 2.31, I decided to explore this subject for the first time. My bench measurements from Spring 2012 punctuated how little I knew about RF bypass and I share these notes as something for me and others to build on.

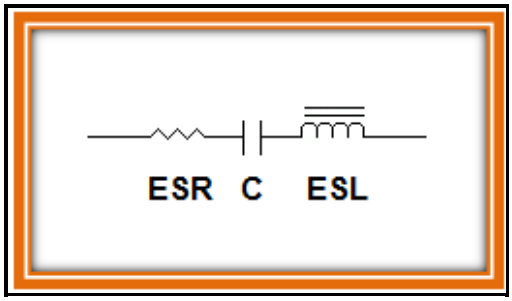
In these experiments, I

1. observed the self resonant frequency of MuRata RPE Series, 50v, 5% capacitors with X7R temp compensation at 0.1, 0.01 and 0.001 µF.

2. examined a wire short, plus 1 and then 2 Johanson Dielectric 0.01µF, 50v, X7R, size 1206 chip capacitors.

3. tested a 0.1 µF RF cap plus a parallel 2.2 µF electrolytic capacitor to look at parallel resonance side effects.

4. attempted to reduce the Q of some parallel capacitors to reduce unwanted high impedance peaks.



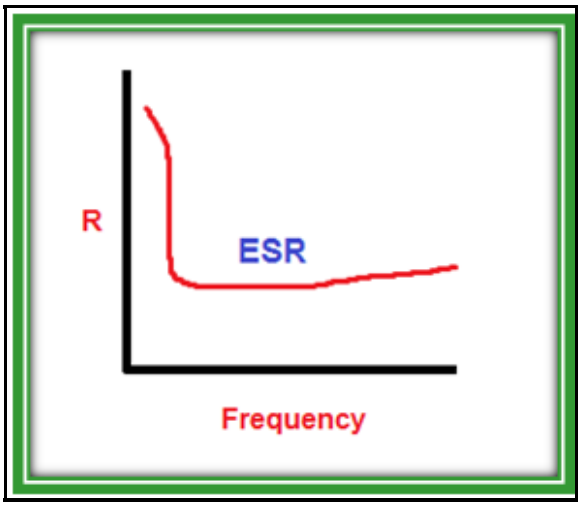
Above — The frequency dependent components of a capacitor are shown in this capacitor equivalent circuit schematic; essentially an RLC network. Engineers use mathematical formulae to describe the components of a capacitor along with reactance and with this math, you might derive an unknown variable from available data so it's worth diving into on your own.

ESR or equivalent series resistance = the sum of all of a capacitors' resistive components. Expressed in ohms, ESR acts like a resistor in series with the capacitor. Normally we desire capacitors with an ESR as low as possible. Consider reading the capacitor datasheets for those your stock and/or searching for information regarding low ESR capacitors on the Internet.

ESL refers to the equivalent series inductance; the sum of all the capacitor's inductive components. This includes lead length in hole-through parts.

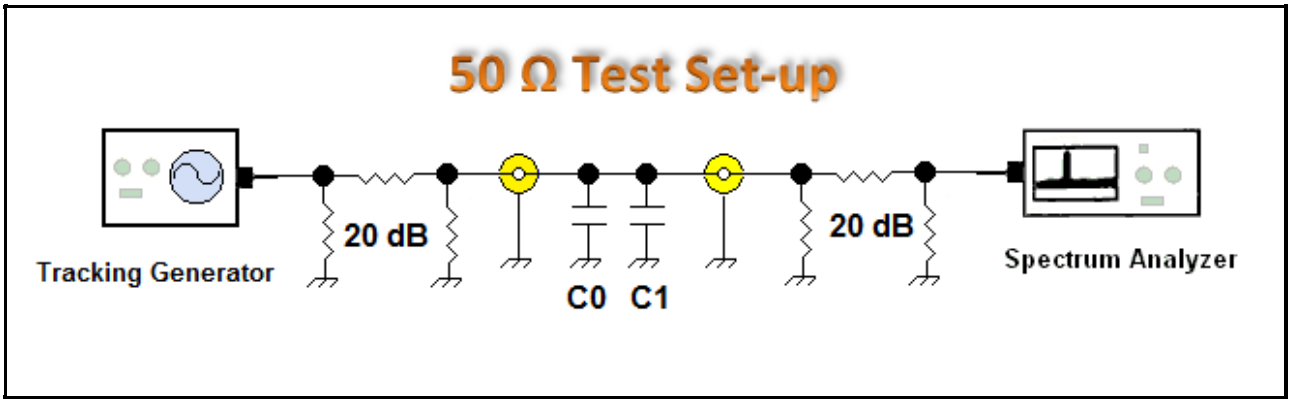
In a given capacitor, the series resonant frequency is the frequency where the inductive reactance from the ESL = the capacitive reactance, but since the 2 reactances are 180 degrees opposite in phase, they cancel to drop the impedance to 0 and the capacitor acts like a resistor at its ESR.

The series inductance of a capacitor may be determined using a network analyzer and unfortunately this in unattainable by most average builders. When designing RF bypass with network analysis, we strive for a low impedance over a wide frequency range, although small ripples typically occur.



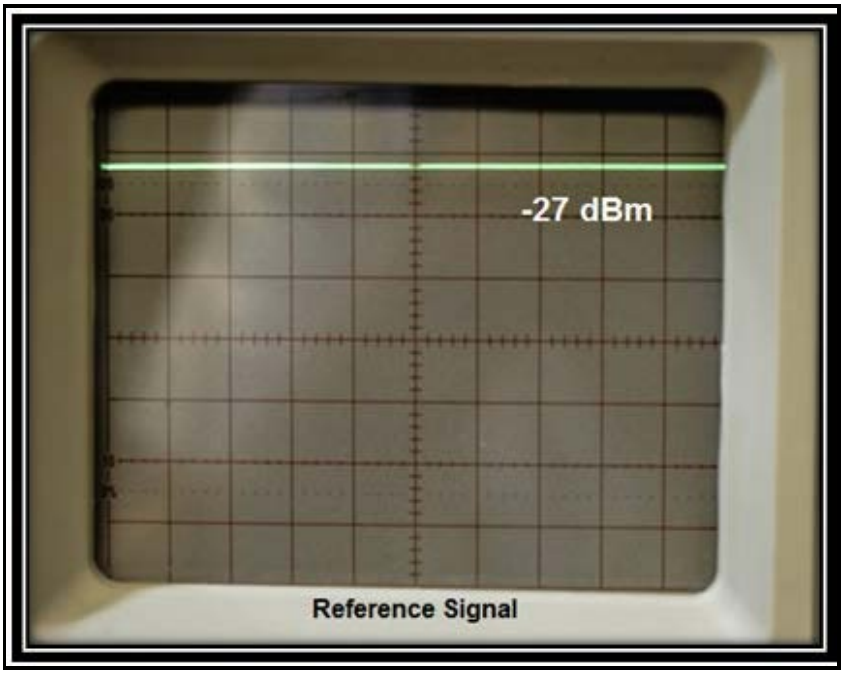
Above — A plot of equivalent series inductance. ESR tends to increase with frequency.

Methods



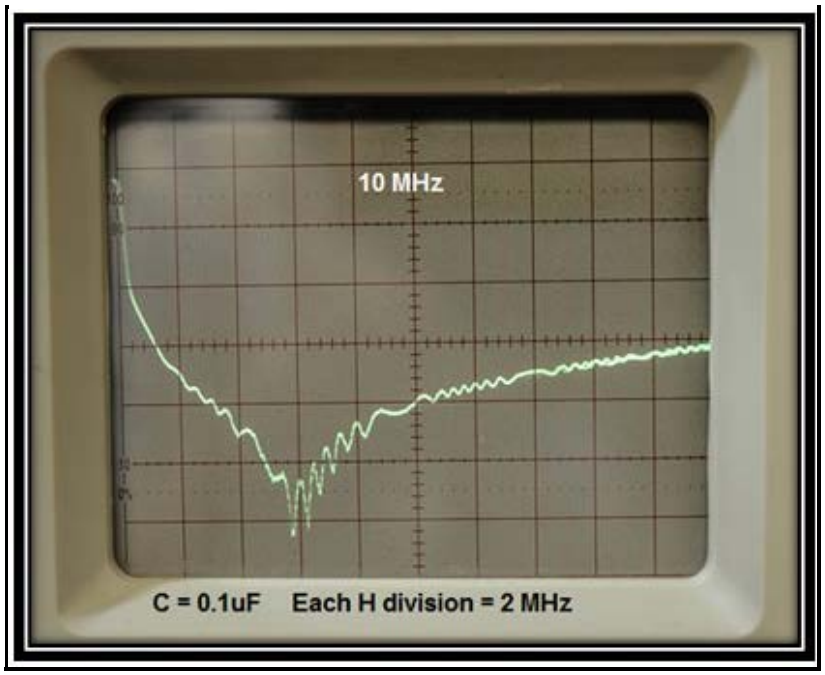
Above — My test set up. I performed all analysis with a tracking generator plus spectrum analyzer. The 50 Ω system used short coax patch cables fitted with BNC connectors with 20 dB attenuator pads before and after the capacitors under test. The capacitors shown as C0 and C1 were soldered on a copper board with short leads and BNC connectors. C1 is omitted when evaluating only 1 capacitor.

You may also perform capacitor self resonant frequency testing with a vector network analyzer, a signal generator plus a 50 Ω terminated scope, or with a sweep generator ramp-driving the oscilloscope X input while simultaneously driving a VCO with logarithmic output to the Y oscilloscope input. SPICE simulations may also yield insight.

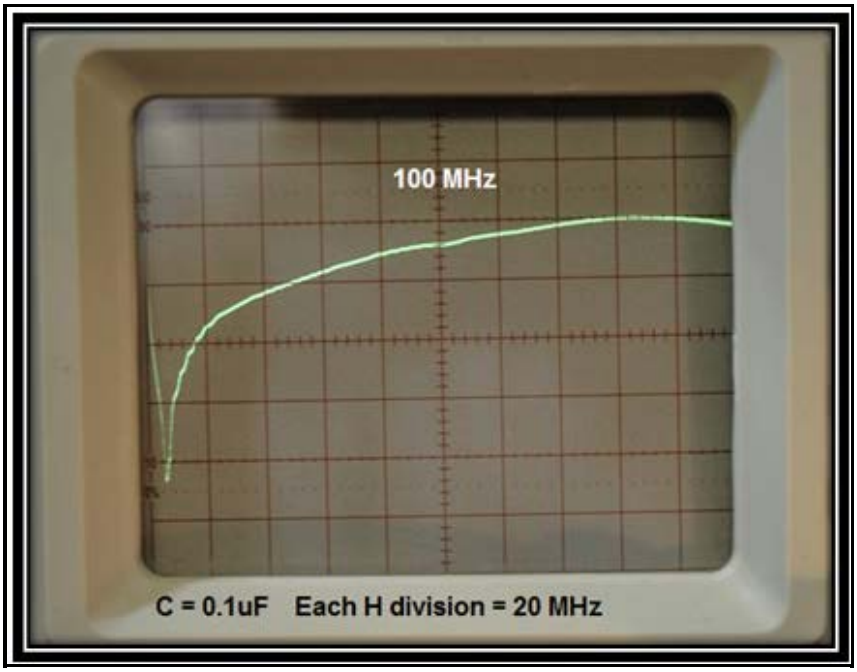


Above — The -27 dBm reference with a through-connector between my 2 coax patch cables (C0 + C1 board removed).
To save time I shot these SA photos handheld and prefer a slower shutter to capture a nice CRT tracing, so some of the photos show a little hand jitter.

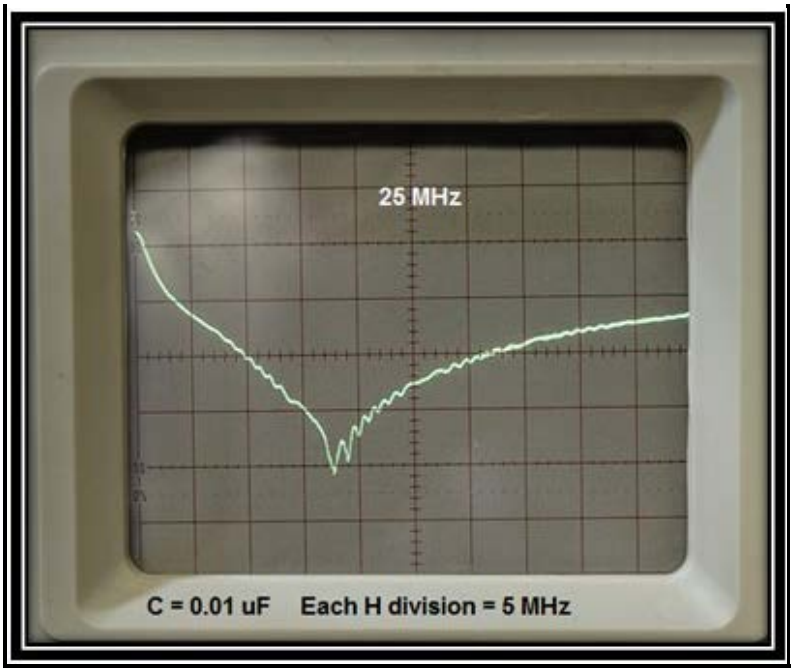
Single Shunt Capacitors



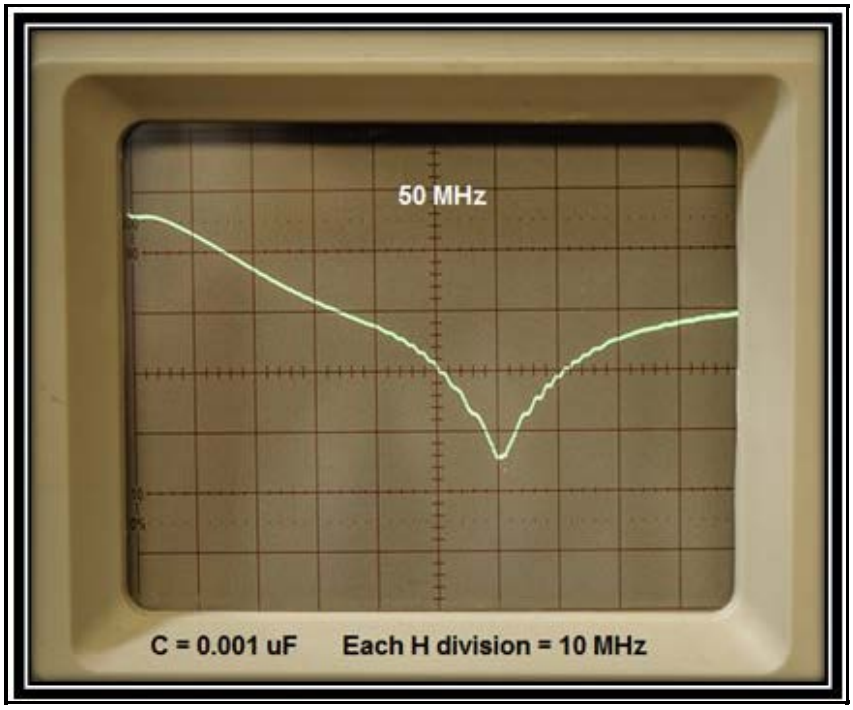
Above — C0 = 0.1 µF. I view the capacitor like a trap. At almost 5.8 MHz lies the peak attenuation, or lowest impedance — this is C0's self resonant frequency. The peak bypass frequency lies ~ 60 dB down. At 20 MHz, the attenuation is only ~ 30 dB.



Above — Another shot of the 0.1 µF bypass cap with a 200 MHz span. At 50 MHz, the reference signal lies only ~ 17 dB down. At 100 MHz, the attenuation is only ~ 11 dB — this hardly qualifies as “bypass” much above the self resonant frequency. Above the self resonant frequency, a capacitor's XL affects impedance more than the ESR and XC of the capacitor.



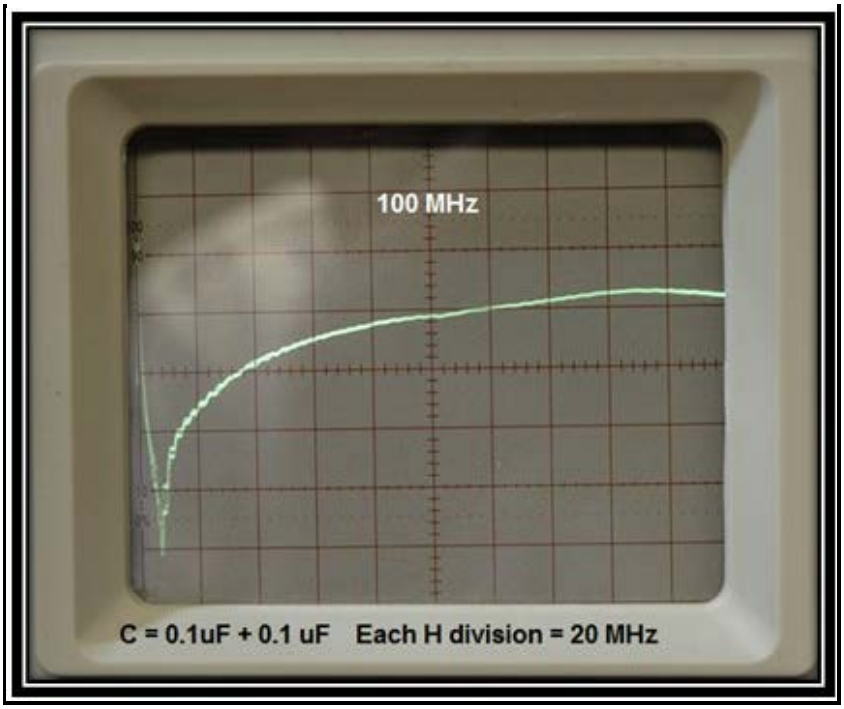
Above — C0 = 0.01 µF. The peak bypass frequency (capacitor self resonant frequency) is centered at 17.5 MHz and is ~ 50 dB down; not as deep as with the 0.1 µf cap. At 50 MHz, the attenuation is ~ 21 dB.



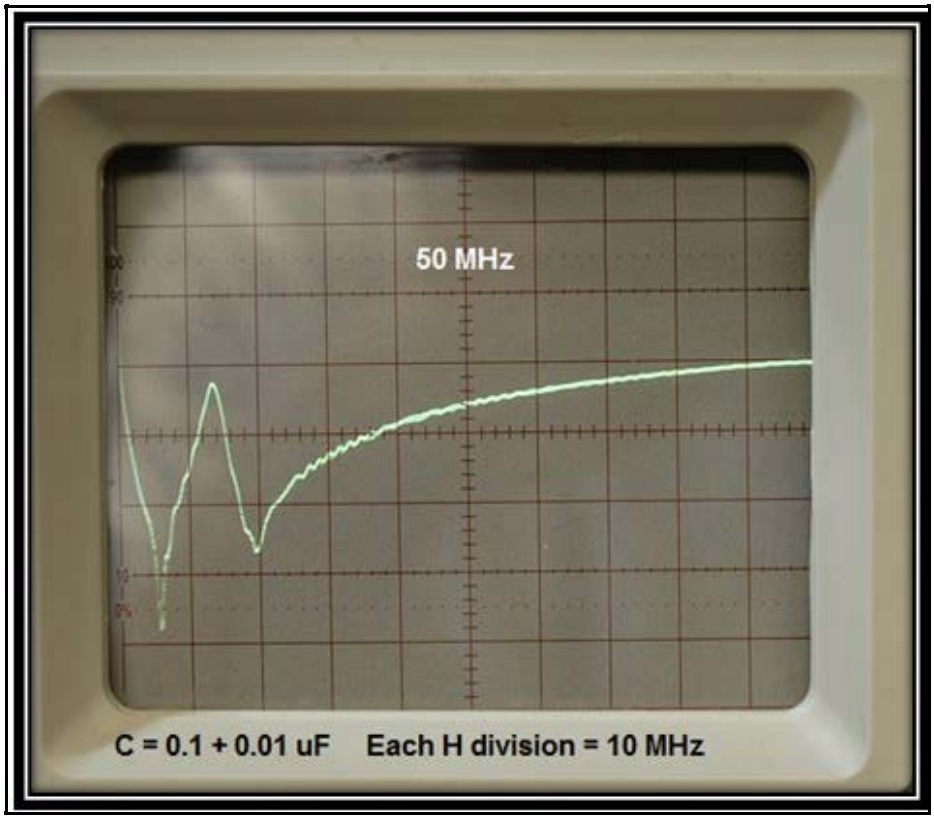
Above — C0 = 0.001. The response is peaked at 62 MHz with an attenuation of ~42 dB. At 100 MHz, the signal is 17 dB down. Again, the peak attenuation looks diminished compared to that of the 0.1 µF and the 0.01µF caps.

Capacitors in Parallel

Now I placed 2 caps in parallel (C0 + C1) as some builders do to try and garner a wider attenuation bandwidth.

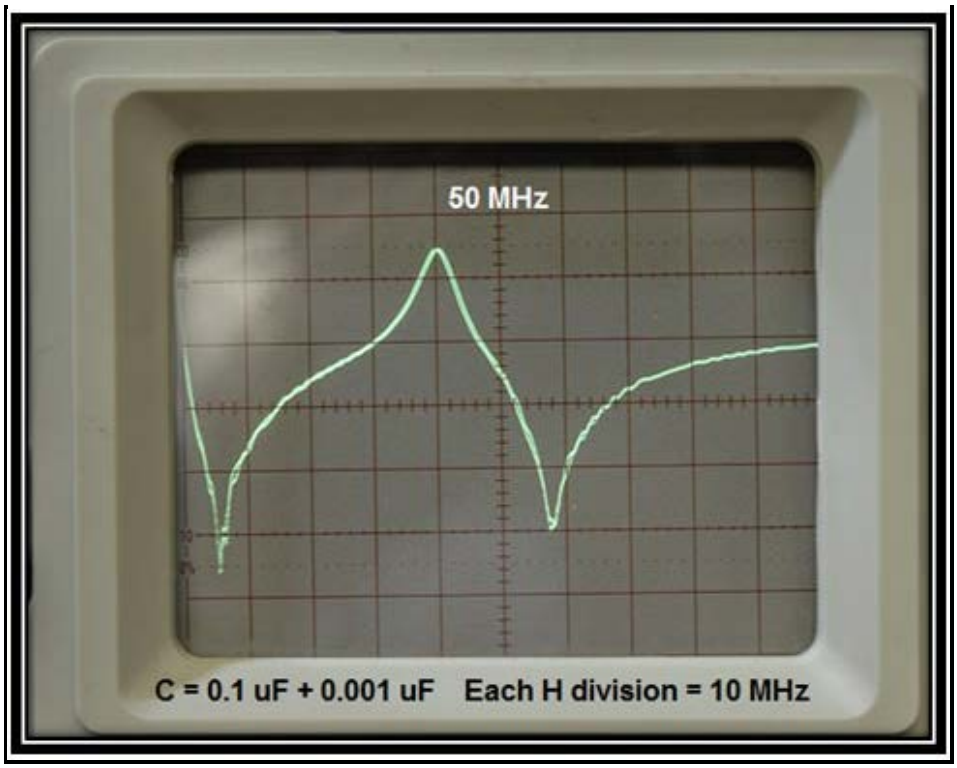


Above — C0 = 0.1 µF + C1 = 0.1 µF. The peak attenuation = 60 dB at 8 MHz; up 2 MHz from that of the single 0.1 µf capacitor. At 100 MHz, attenuation = ~ 19 dB — better than a single 0.1 µF but still low.



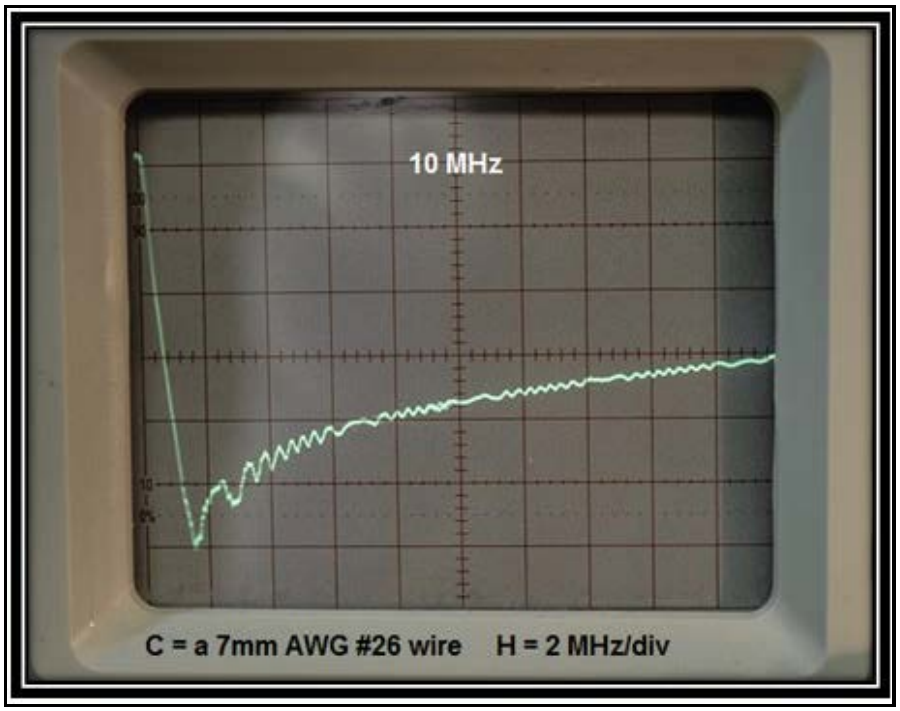
Above — C0 = 0.1 µF + C1 = 0.01 µF. Yikes! With the 2 different cap values, we get an unfortunate high impedance blip peaking at 13 MHz. Each capacitor exerts its self resonant frequency, but in between these self resonant frequencies, lies a disaster.
When placed in parallel, the inductance of 1 capacitor resonates with the capacitance of the other to form a parallel resonance — leading to a high impedance — that blocks RF bypass and peaks at a specific frequency.

But wait. Things can get worse:

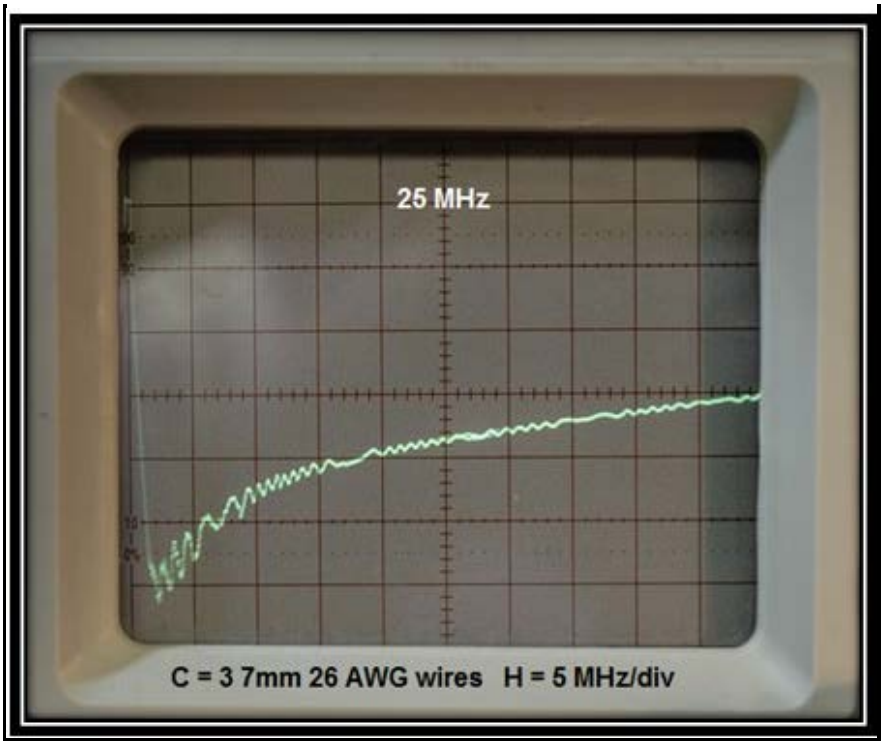


Above — C0 = 0.1 µF + C1 = 0.001 µF. The wide value variance between these 2 capacitors creates a huge, high impedance spike where the attenuation is only about 6 dB at 40 MHz. Catastrophic bypass indeed. катастрофа.

A 7 mm Length of Copper Wire



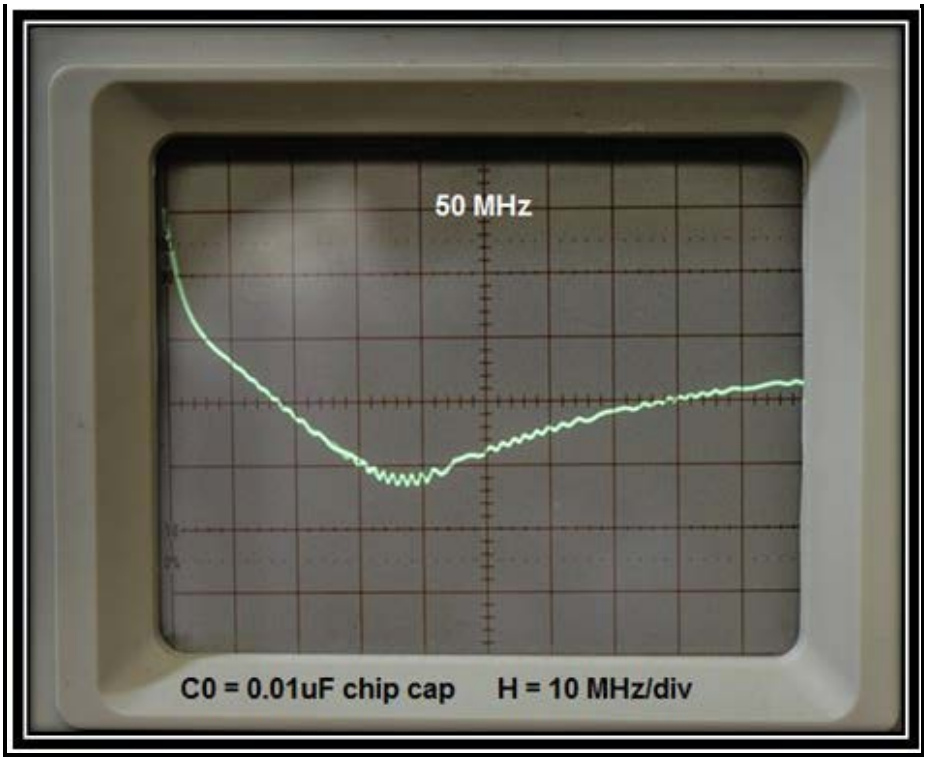
Above — A 7 mm piece of 26 gauge copper wire was shorted to ground instead of C0. This wire measured at ~7 nH of inductance and I saw that attenuation decreases with frequency from 62 dB at 8 MHz to ~33 dB at 20 MHz. Even a short piece of wire doesn't exhibit a flat, wideband bypass.



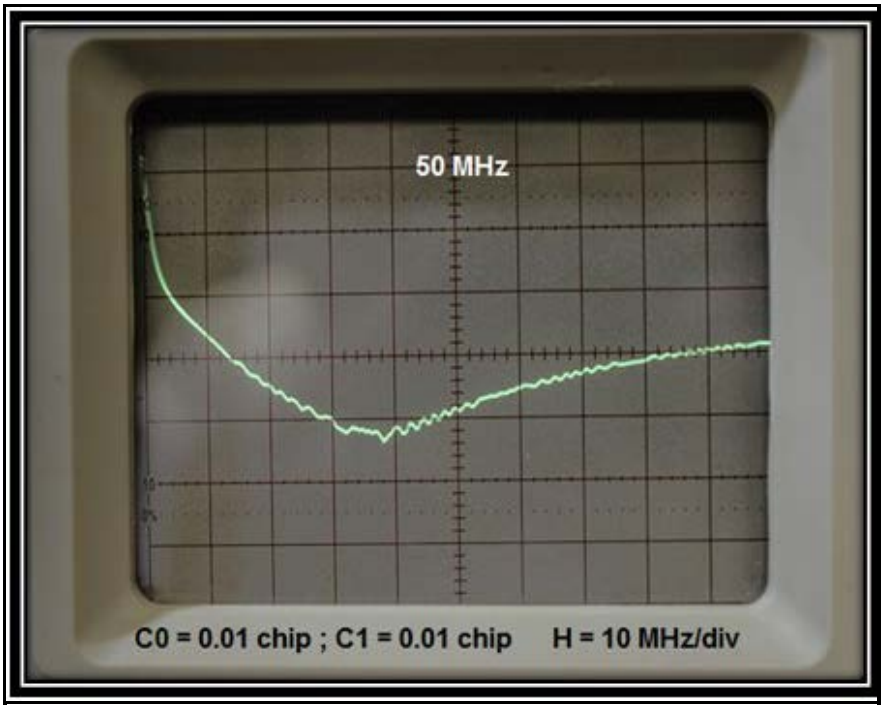
Above — This spectrum analysis shows three 7 mm wires shunted to ground – not much different than 1 wire.

0.01 µF Chip Capacitor(s)

Above — The magnified copper board that I tested one or two 0.01 µF chip capacitors. You can see 1 capacitor soldered in.

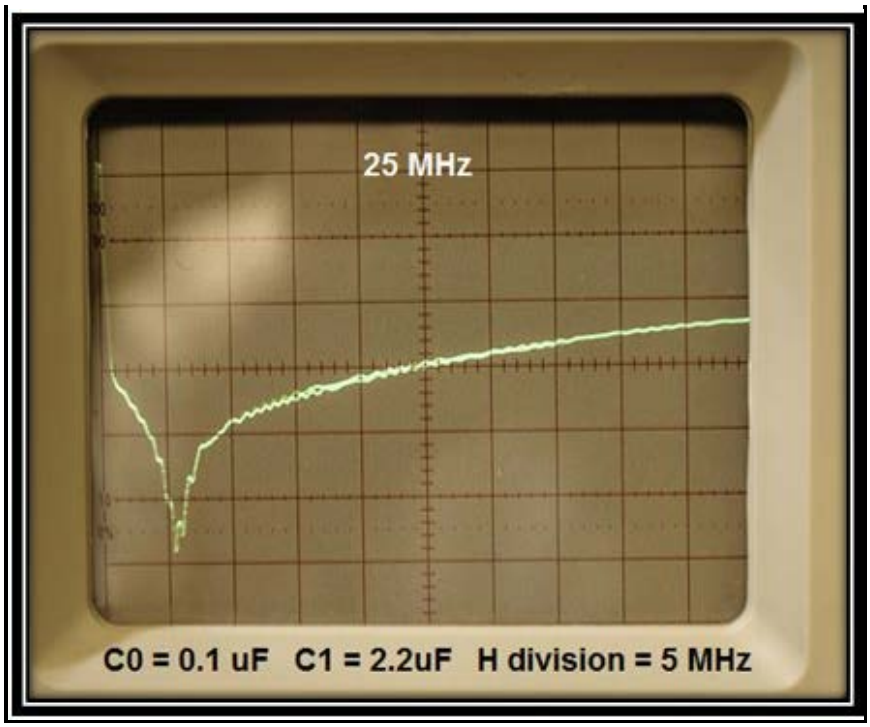


Above — C0 = 0.01 µF SMT cap. The SMT parts exhibited a peak attenuation of 45 dB at ~37 MHz. The attenuation dip lacks the sharp peak of the hole-through 0.01 µF cap shown eariler and exhibits a somewhat wider bandwidth. The self resonant frequency of the chip capacitor is 5 MHz higher than the particular hole-through capacitor I measured. Click for a side by side photo.



Above — C0 = 0.01 µF + C1 = 0.01 µF. The SMT parts exhibited a peak attenuation of 45 dB at ~37 MHz; similar to the single 0.01 µ F chip cap, but with a few more dB attenuation between 10 and 20 MHz.

0.1 µF Ceramic + a 2.2 µF Electrolytic Capacitor



Above — C0 = 0.1 µF and C1 = 2.2 µF. The low Q 2.2 µF C1 electrolytic cap did not create the a parallel resonance with C0. Shaky photo — sorry. I also tested a 10 and 22 µF cap in parallel with C0 and saw no disturbance caused by a parallel resonance between a big AF capacitor and C0 (an RF value cap) with my RF spectrum analyzer.

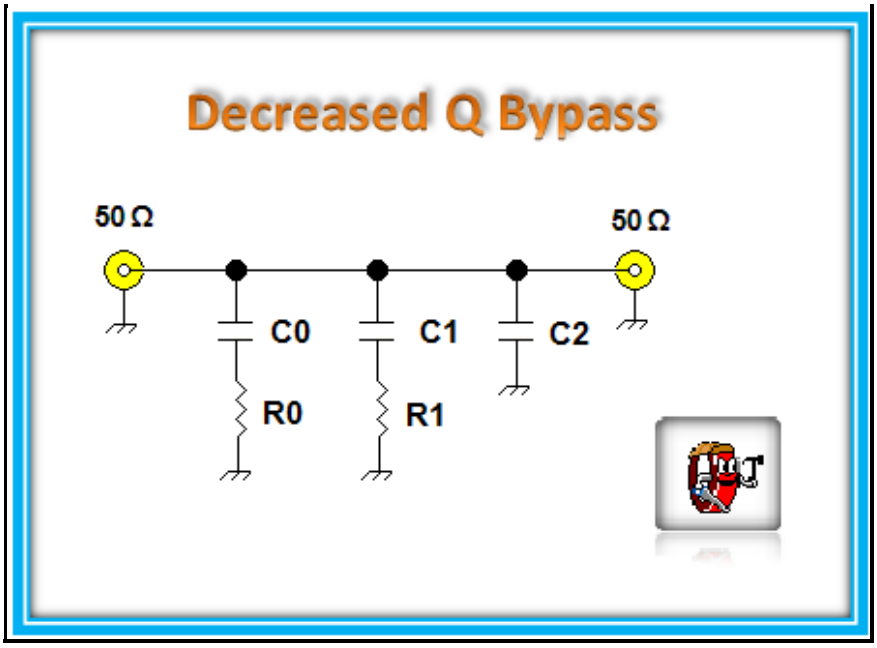
Some additional experiments applying a low Q *A*F capacitor plus a ceramic *RF* capacitor for wideband bypass yielded some interesting results and I'll present these in a future project.

Capacitors in Parallel with a Series Resistor to Lower Q

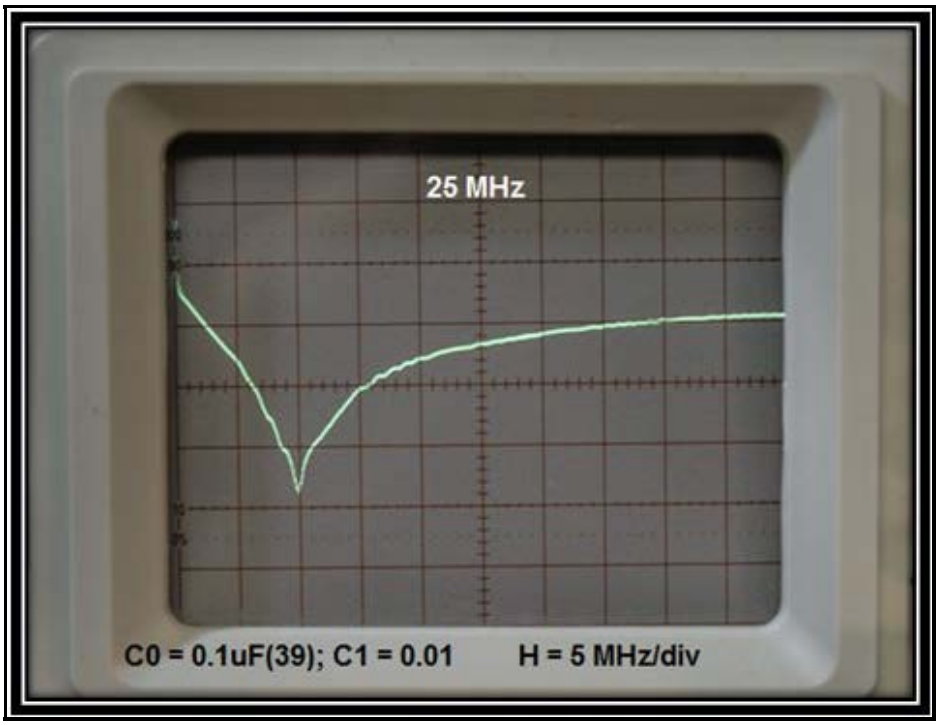
In previous experiments, placing 2 RF capacitors in parallel led to the formation of a peaked high impedance blip between the low impedance peaks set by the self-resonant frequency of the 2 capacitors. If multiple capacitors are soldered in parallel, the series inductance of each capacitor will resonate with the capacitance of the next smaller C value.

One solution is to put a resistance in series with all but 1 of the parallel capacitors so that the Q of resonance formed by this capacitor's series inductance and the capacitance of the next smaller capacitor is low. If capacitors exhibited 0 inductance then putting capacitors in parallel would be fine, however, since capacitors exhibit inductance, a parallel resonant frequency may occur with capacitors in parallel.

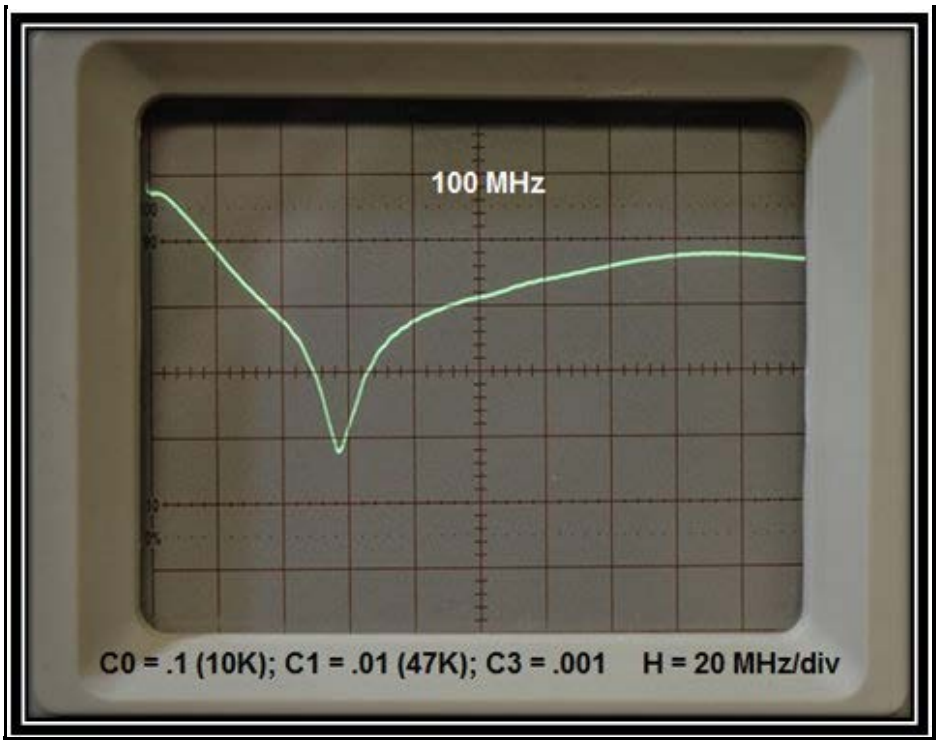
I found applying a series resistance to lower Q may flatten the impedance versus frequency response of the bypass network, but didn't decrease the impedance at any 1 frequency. Optimal bypassing or achieving the lowest impedance over a wide frequency range presents a complex topic that might even challenge some engineers.



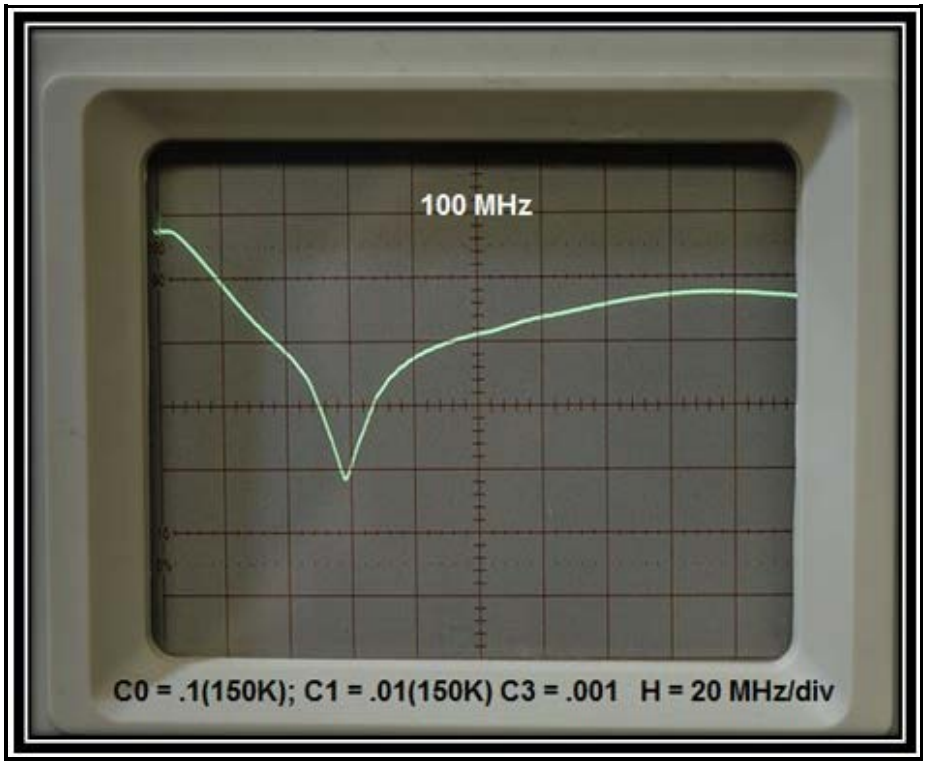
Above — A method to exact wideband bypass.



Above — My first try with C0 = 0.1 µF, R0 = 39 Ω and C1 = 0.01 µF. I arbitrarily placed the 39 Ω resistor in the R0 slot and saw that the high impedance peak seen earlier disappeared. This gave me the confidence to try 3 capacitors. I had no idea what R value to use and really just wanted to see what happens.

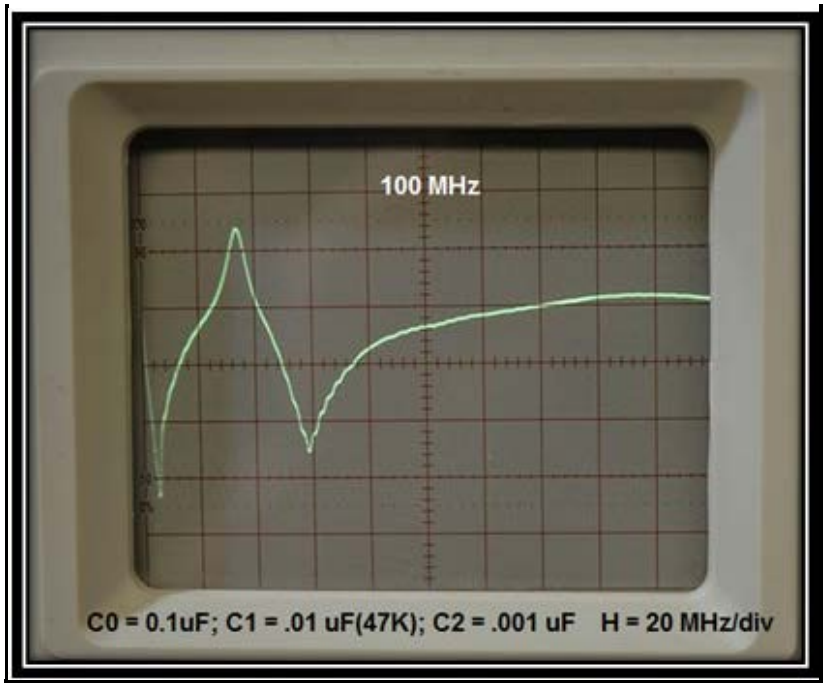


Above — The spectrograph with C0 = 0.1 µF, R0 = 10K, C1 = 0.01 µF, R1 = 47K and C3 = 0.001 µF.



Above — C0 = 0.1 µF, R0 = 10K, C1 = 0.01 µF, R1 = 47K and C3 = 0.001 µF. Again, no high impedance peak response; the self resonant frequency is close to that measured with a single 0.001 uF earlier, however, the peak bypass frequency moved from to 57 MHz from 62 MHz. Changing the resistor values moved the self-resonant frequency and the peak attenuation value a little, but I fell kilometers short of setting a wide band bypass. My approach lacks any real science and I need to step it up.

I hope to learn what capacitor values and types, plus R values to apply. This sounds like a job for simulation as well as further on-bench experiments? After writing this material, I learned that Ken Kuhn wrote an Excel spreadsheet to examine the net impedance of up to 3 capacitors in parallel Click to download.



Above — Just as a gag, I removed R0 from capacitor C0 in 1 circuit and then hooked up the board. The high impedance peak re-emerged.

A Commerical Example

I found a wideband MMIC employing (R1 + C1 and C2) as part of a bypass strategy. Cick for the datasheet excerpt. Note the size of the SMD capacitors; 0603 — tiny caps! My experiments showed some high gain MMICs require careful low inductance grounding and correct part choices or crippling oscillations and other bypass issues might arise.

My Learnings

When we think bypass, we really should think frequency dependent attenuation. The bypass cap is actually a network where impedance versus frequency varies significantly. At its self resonant frequency, a capacitor will exhibit the lowest possible impedance making a single capacitor a relatively narrow-band bypass device. Intuitively, we might want to choose a capacitor with a series resonant frequency at the frequency we wish to bypass, however, if we require a wideband bypass, the need to evaluate our bypass capacitor(s) increases.

In short, above the series resonant frequency of a capacitor, its bypass is basically useless and we should likely ensure that the self-resonant frequency of the particular capacitor we're using is above the highest frequency to be bypassed.

Bypassing with 2 or more unmatched RF caps will lead to an attenuation gap with peak(s) determined by the parallel resonance of these capacitors. Going above a 10:1 capacitor ratio, for example, greater than a 0.1 and a 0.01 µf, may cause a **severe** gap in attenuation at the parallel resonant frequency generated by the 2 capacitors.

Mine and work from more reputable authors clearly shows we should avoid applying parallel **RF** bypass capacitors of *different* values unless we apply a Q-reducing resistor to the capacitor(s) in parallel with a given RF bypass capacitor. Please read EMRFD page 2.3 for more information and watch out for abundant folklore concerning RF bypass.

The need for measurement and analysis challenges us; in some cases, you may realize good attenuation in the radio band of interest, while poorly bypassing the frequencies above it and compromise an otherwise good design.

Capacitor lead length may affect self-resonance at RF.

Future Work

It would be awesome to learn more about getting a wide-band bypass. I want to order some low or ultra-low ESR caps and measure them. My MuRata RPE Series caps specify low inductance; low is relative — how low is low? Should we apply chip capacitors for bypass in our critical circuits such as low noise VHF amps or MMICs?. Am I fussing about nothing? Lots of questions that folklore just won't answer.

Per EMRFD page 2.3, bypass is only half the equation — we need to *decouple* + bypass to filter RF from moving along our DC lines and so forth.

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6. Some Experiments with Chokes plus Decouple and Bypass Filters

Introduction

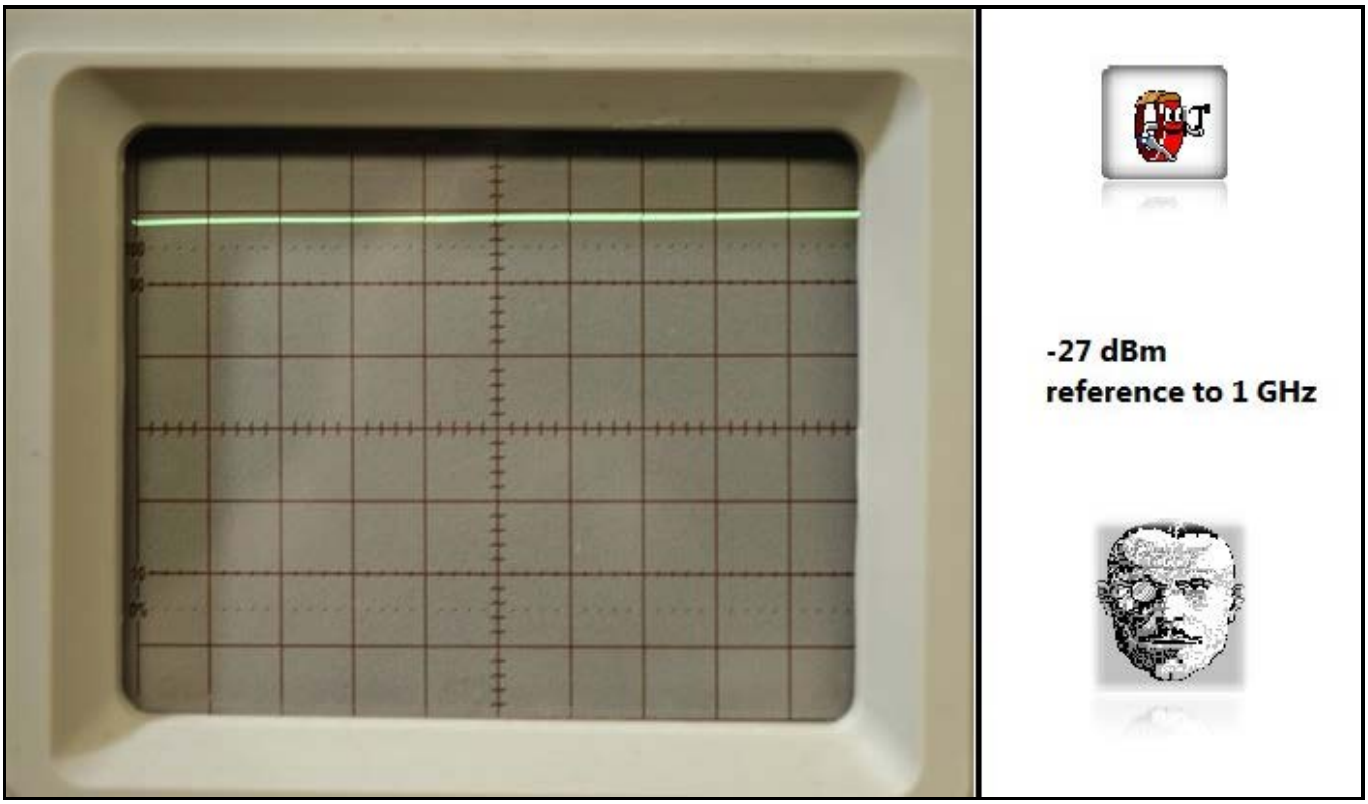
SRF = self resonant frequency; XL = inductive reactance; XC = capacitive reactance. L = inductor; C = capacitor; R = resistor.

Like the capacitor, inductors are networks with R, L and C and possess a SRF. R, L and C may vary with factors including the number of windings, frequency, or whether the L is wound on a ferromagnetic material, or air wound.

Considering R, L and C:

at frequencies below the SRF, XL dominates;
at frequencies above the SRF XC dominates;
at the SRF, the magnitude of XL and XC are equal but 180 degrees out of phase leaving resistance to dominate.

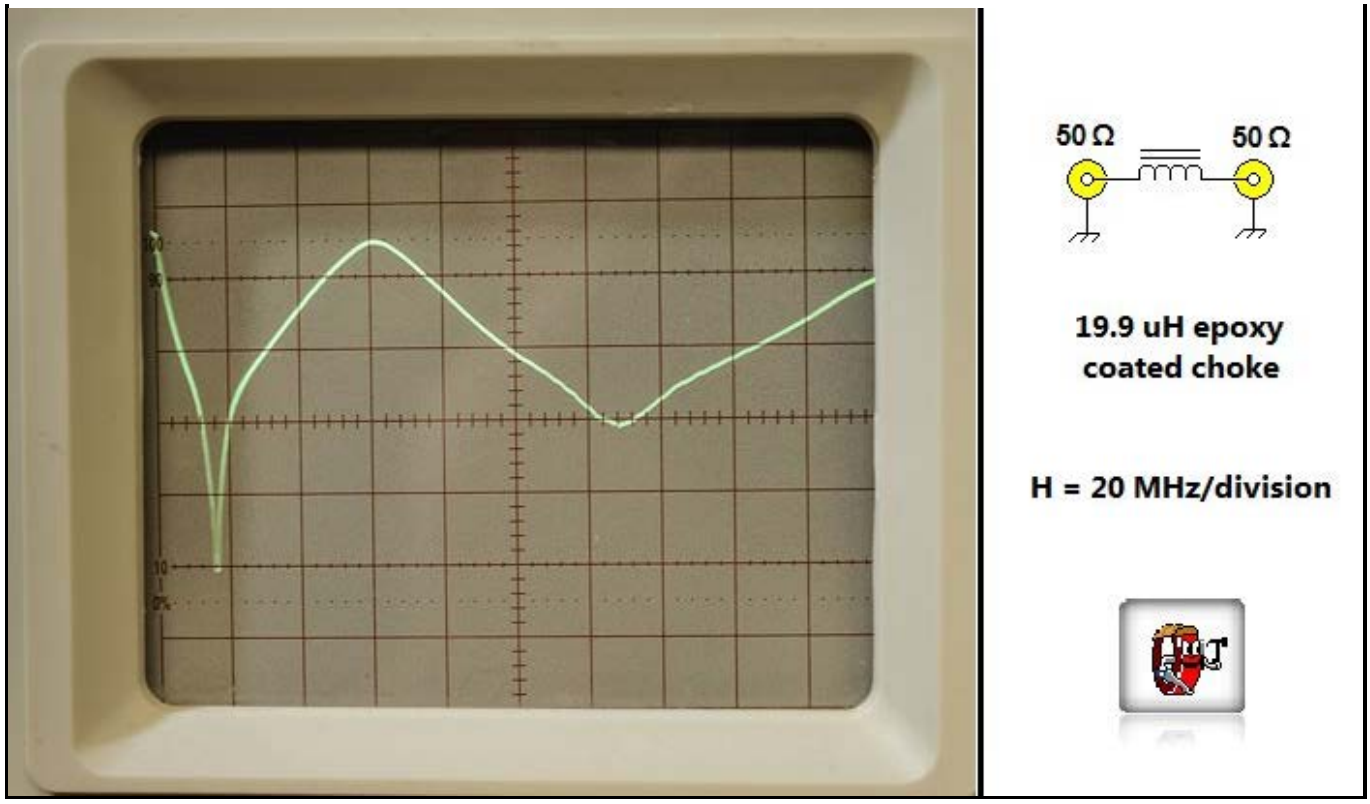
I encourage you to learn more by visiting the fabulous web site of David, G3YNH.



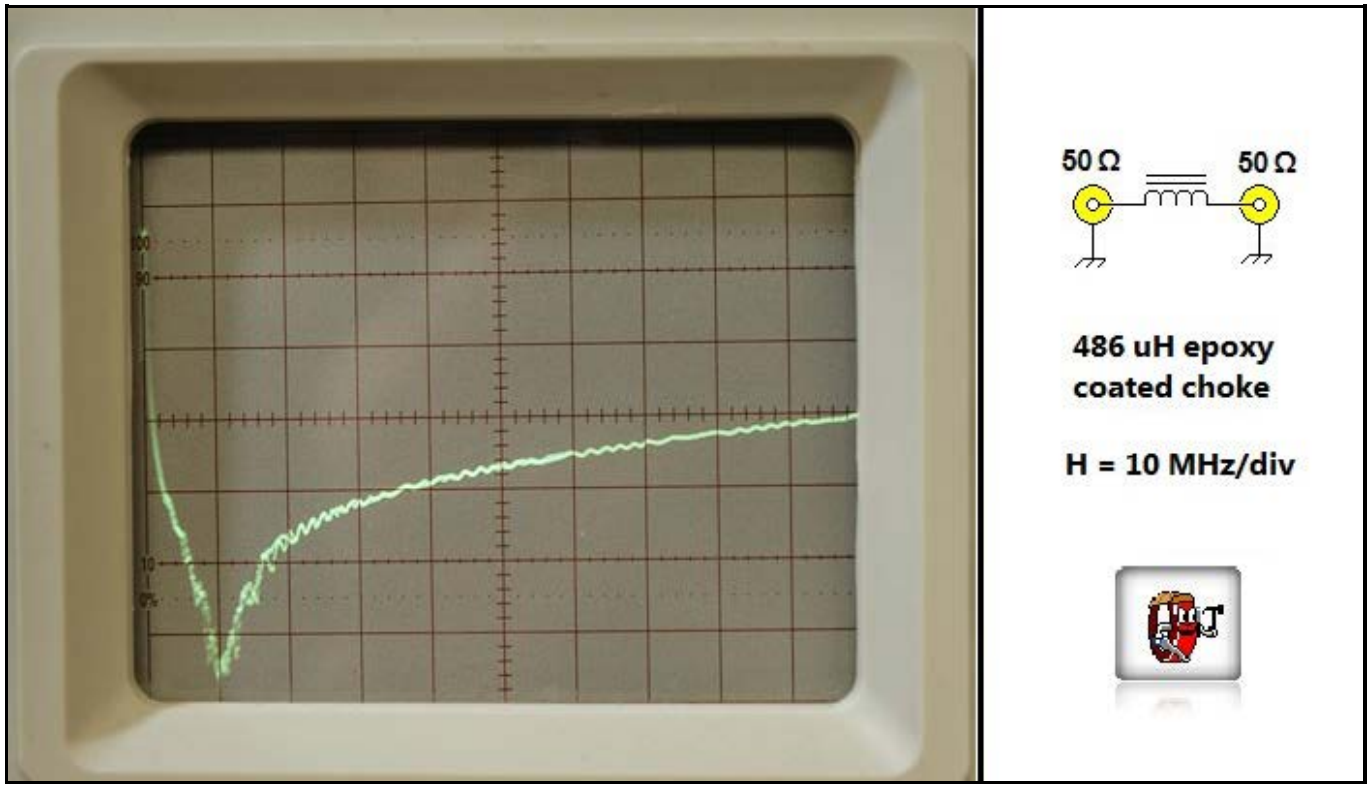
Above — Reference signal at -27 dBm. I used the exact test method shown in Section 5.

For those unaware, the spectrum analyzer screen is divided into 10 by 10 graticules. Each vertical division represent a 10 dB change; read down from the reference -27 dBm to measure the attenuation of the reference signal in dB. Horizontal divisions represent frequency; start at 0 on the left hand side and increment as specified on each figure.

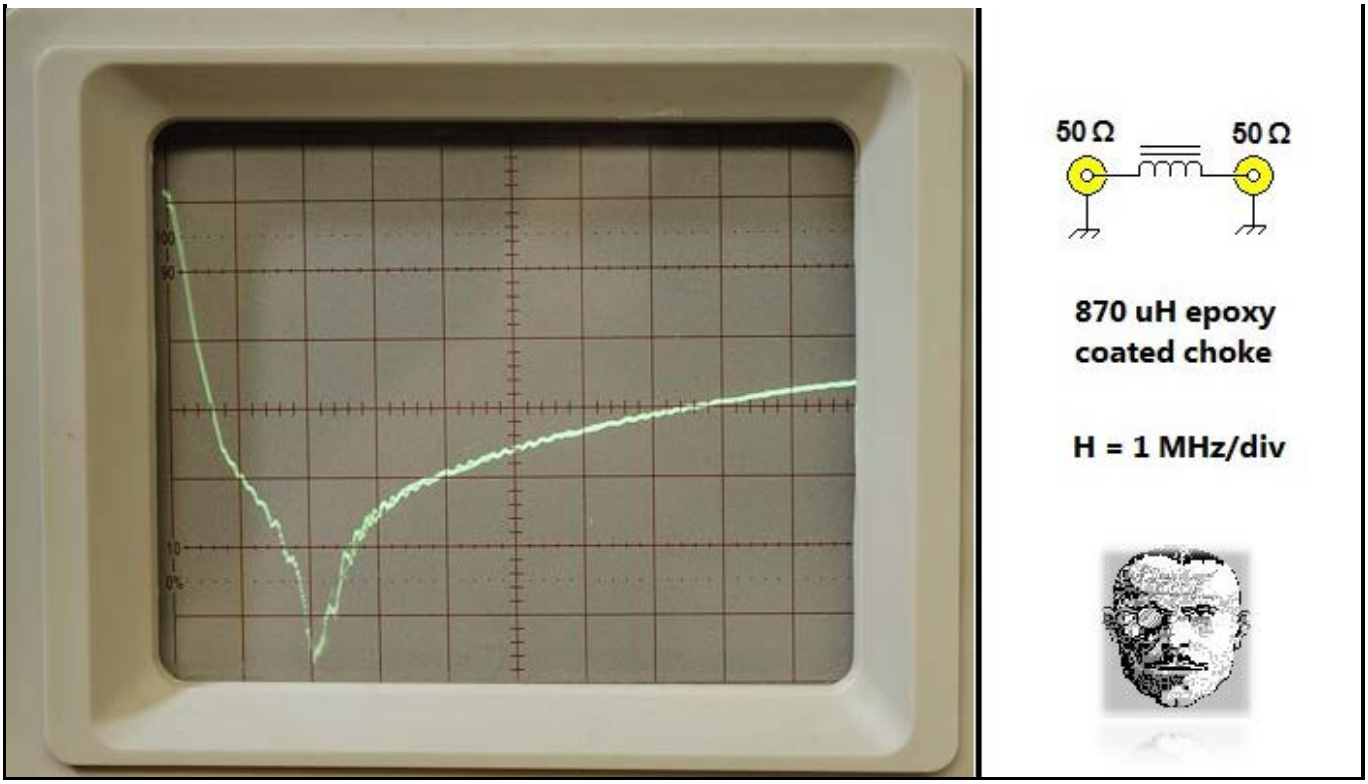
A Few Inductors



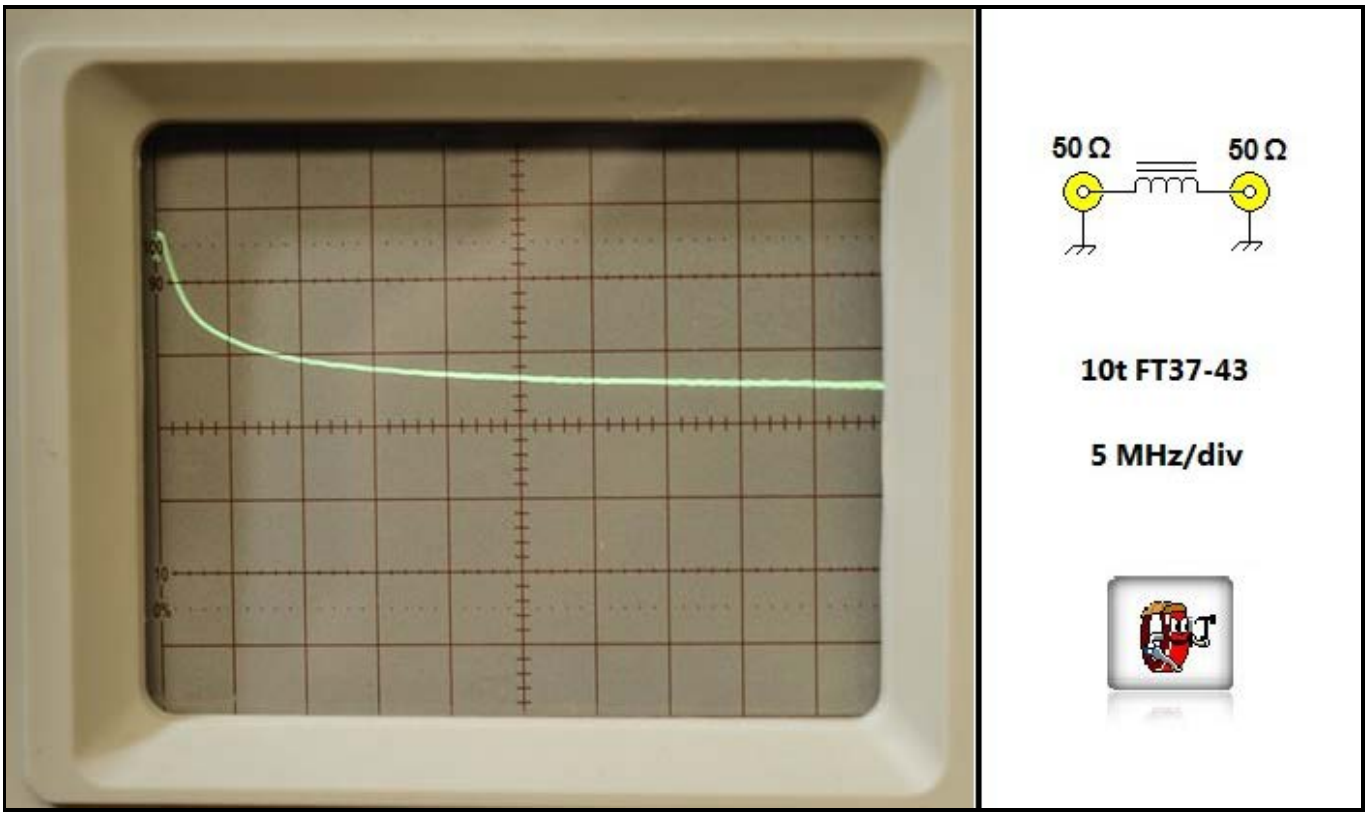
Above — A 19.9 uH epoxy coated choke that exhibits a primary SRF at 18 MHz and a second, smaller SRF at ~128 MHz. This wretched L gave me grief at 63 MHz. After measurement, I tossed it in the garbage can.



Above — A large, junk box choke with an SRF at about 10 MHz.



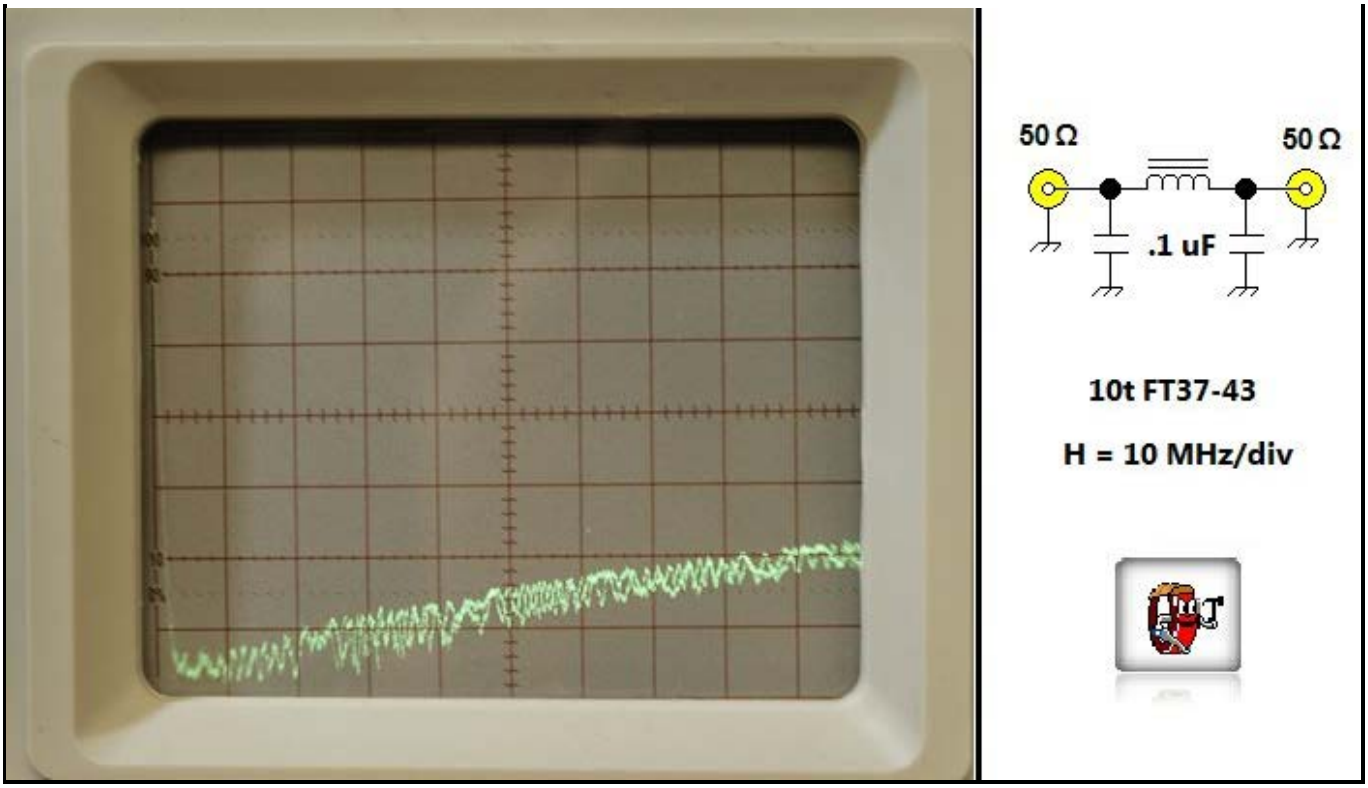
Above — I rarely use these big chokes: 870 µH with a SRF at about 2 MHz.



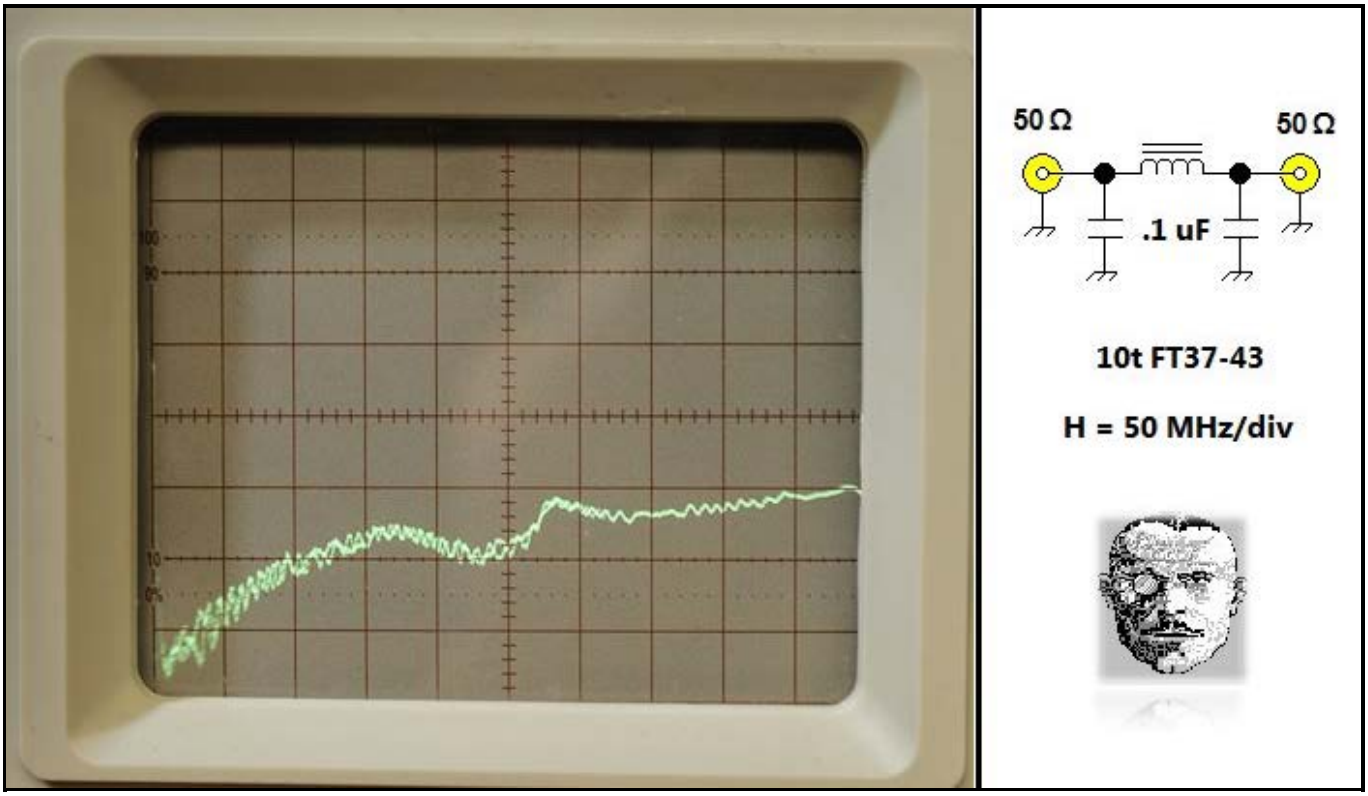
Above — A common L on our benches — 10 turns of #26 AWG on a FT37-43 ferrite toroid.

I couldn't measure the SRF with any span on my spectrum analyzer. I expect that a parasitic capacitance lies in parallel with the inductance, but the #43 material, with its low Q and high losses blankets the usual deep notch we see when the L exhibits a higher Qu.

10 Turns on a FT37- 43 with a Bypass Capacitor Shunting Each End

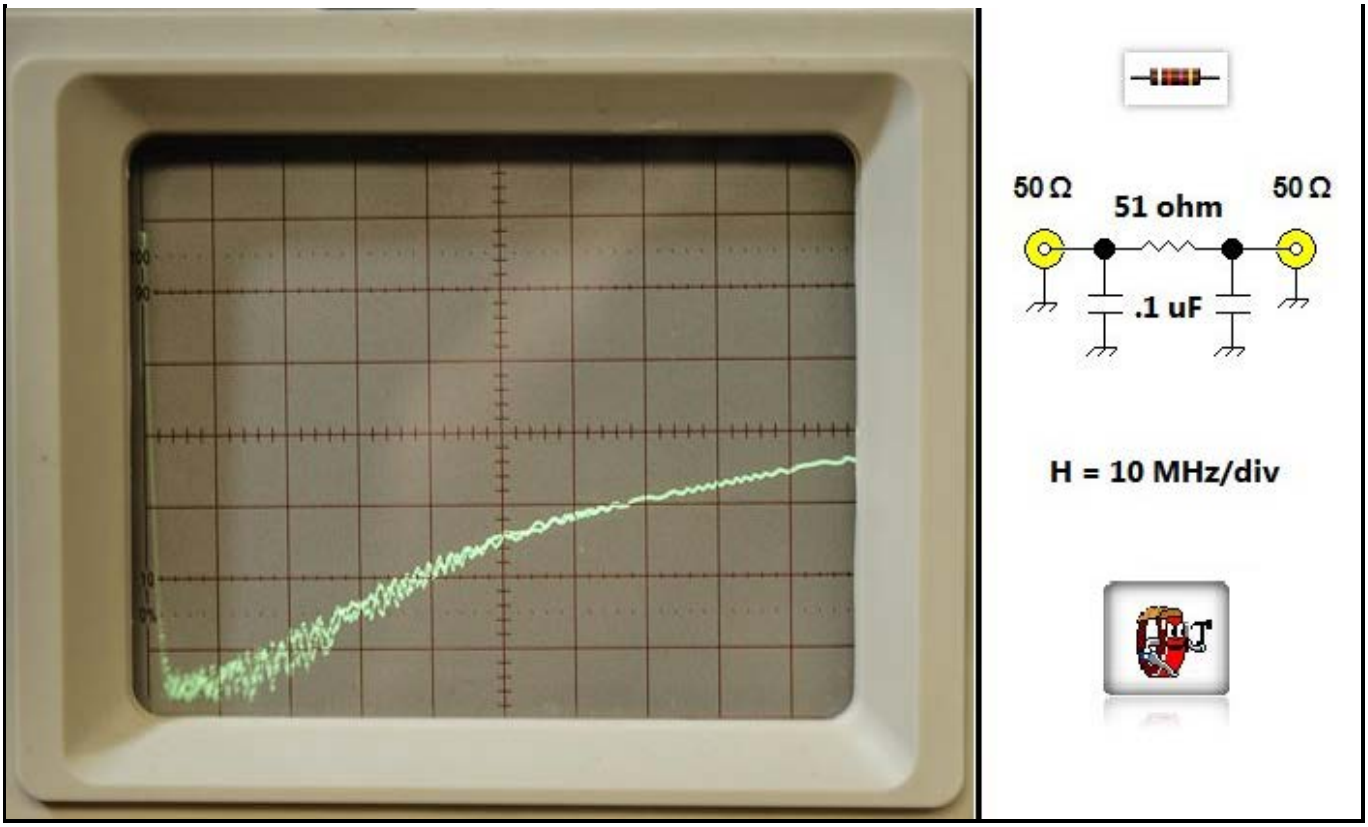


Above — Look at the big difference after adding shunt capacitors to a 10 turn FT37-43! Even at 100 MHz the attenuation lies nearly 50 dB down. Now I understand why Wes says **decouple plus bypass** when filtering our DC lines and so forth.

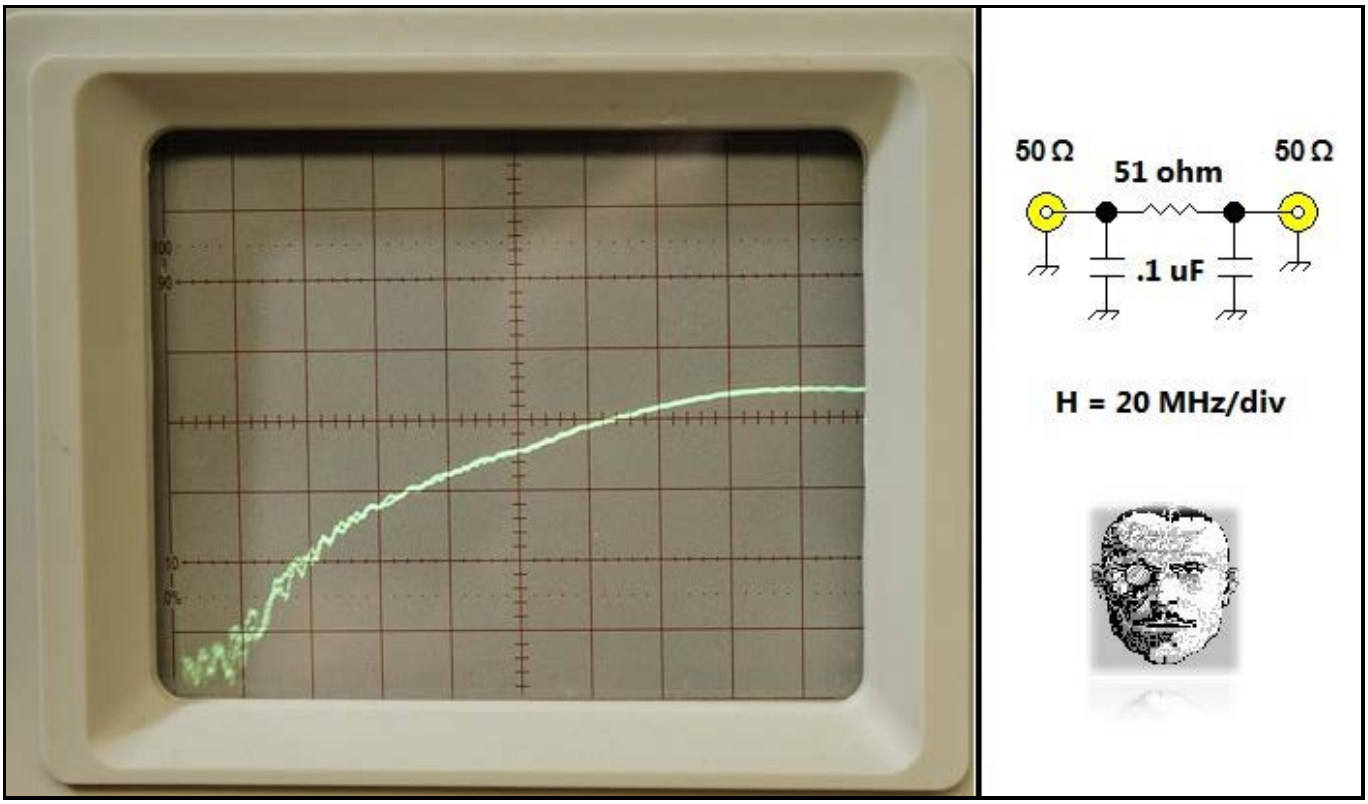


Above — The 10 turn FT37-43 coil with 0.1 µF shunt caps measured out to 500 MHz. Pardon the camera shake; I took all the photos hand held to save time.

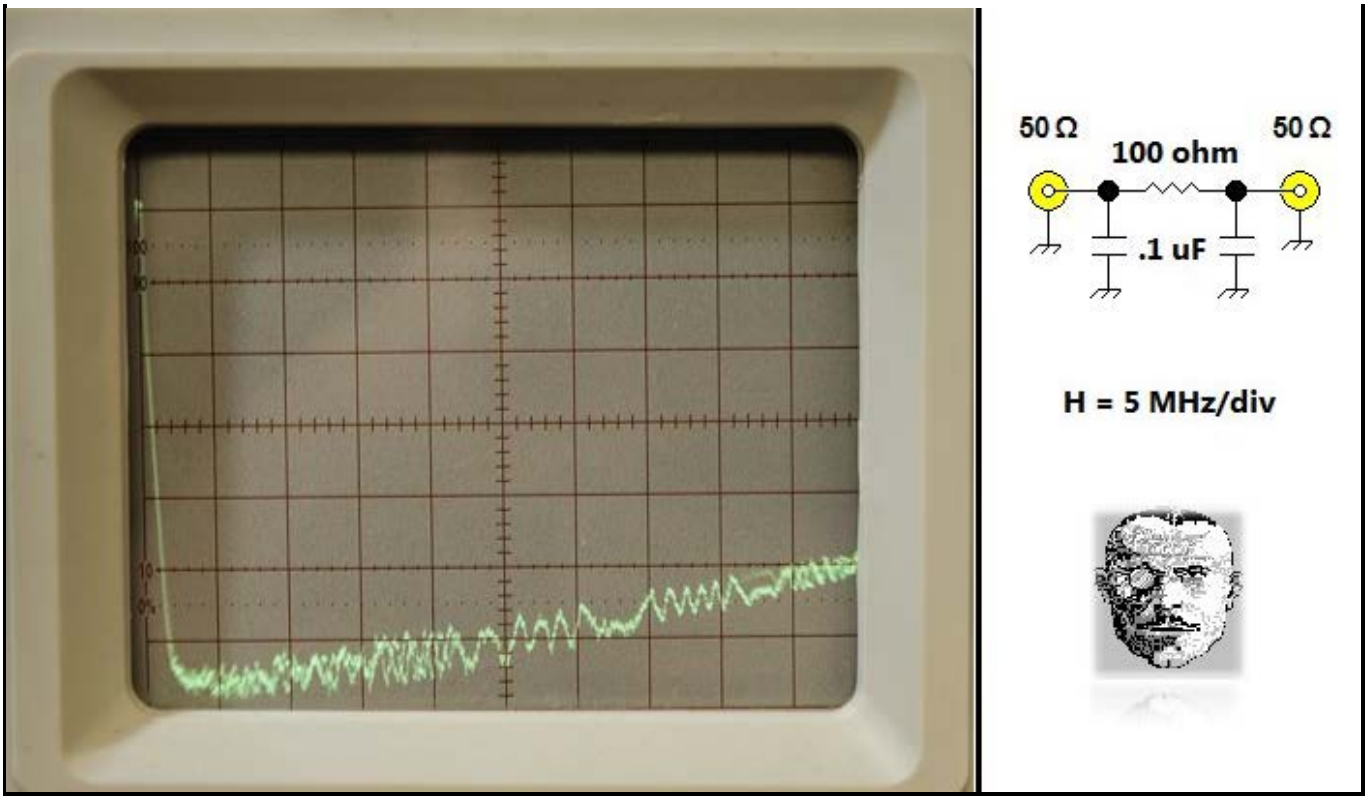
A Resistor with a Bypass Capacitor Shunting Each End



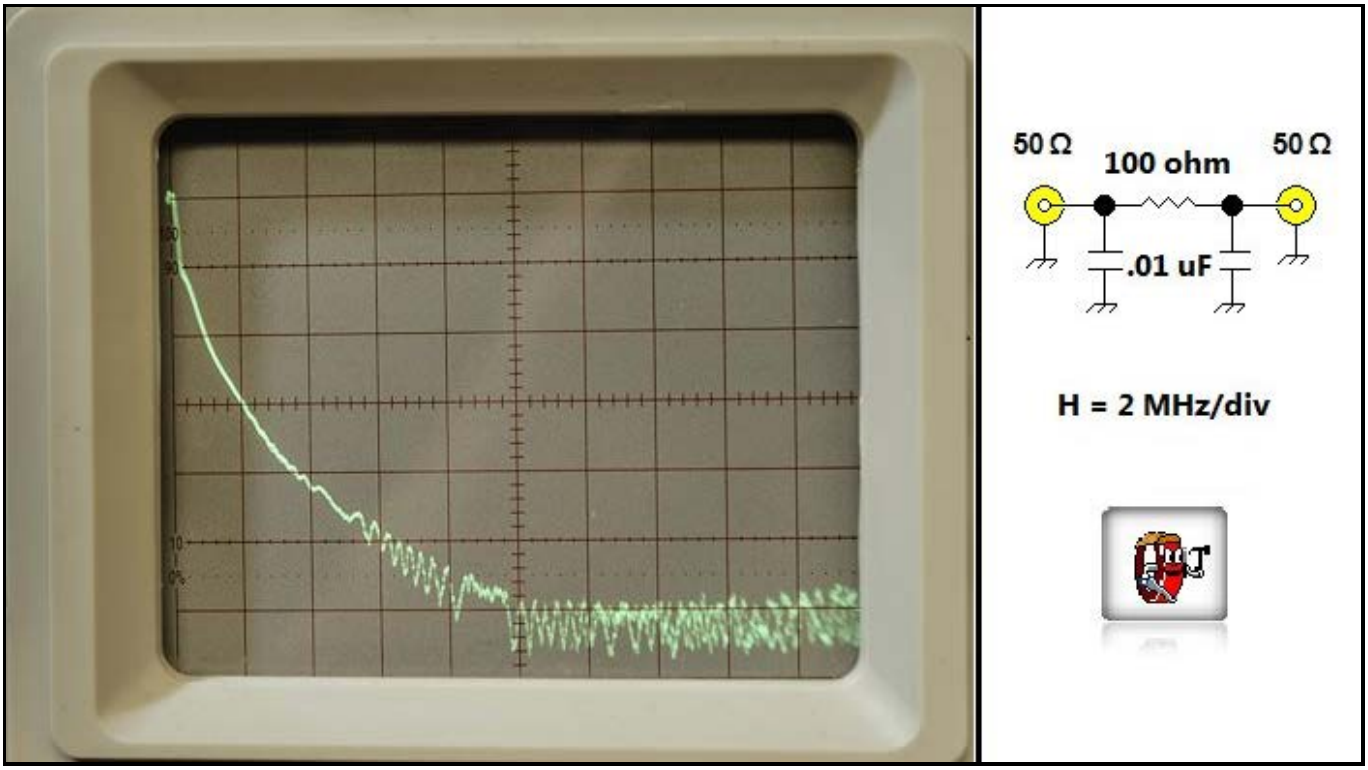
Above — A 51 Ω resistor bypassed with 0.1 µF capacitors at each end spanned out to 100 MHz. Even at 30 MHz, the attenuation looks stellar.



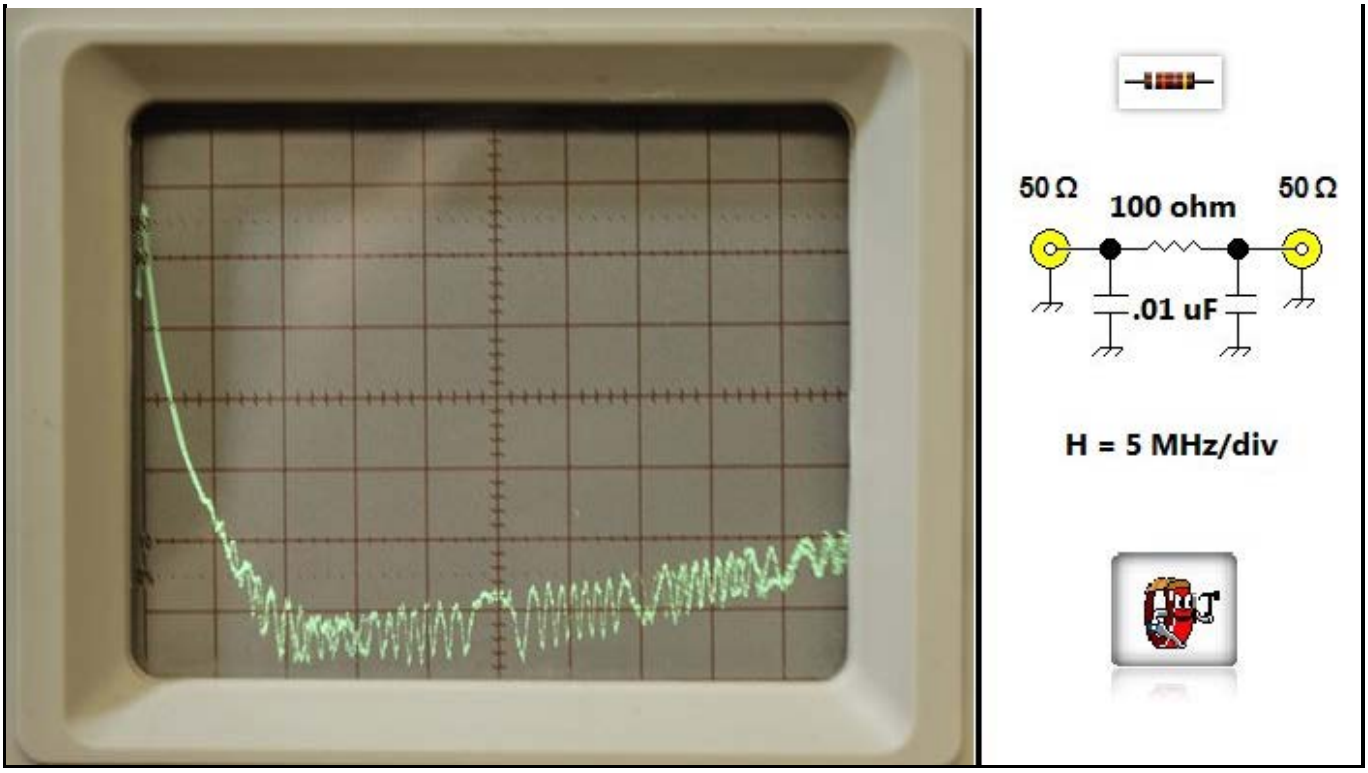
Above — The 0.1 µF bypassed 51 Ω resistor out to 200 MHz. I often use a 51 Ω decoupling resistor with appropriate capacitor values in active circuits that draw from 10 - 18 mA.



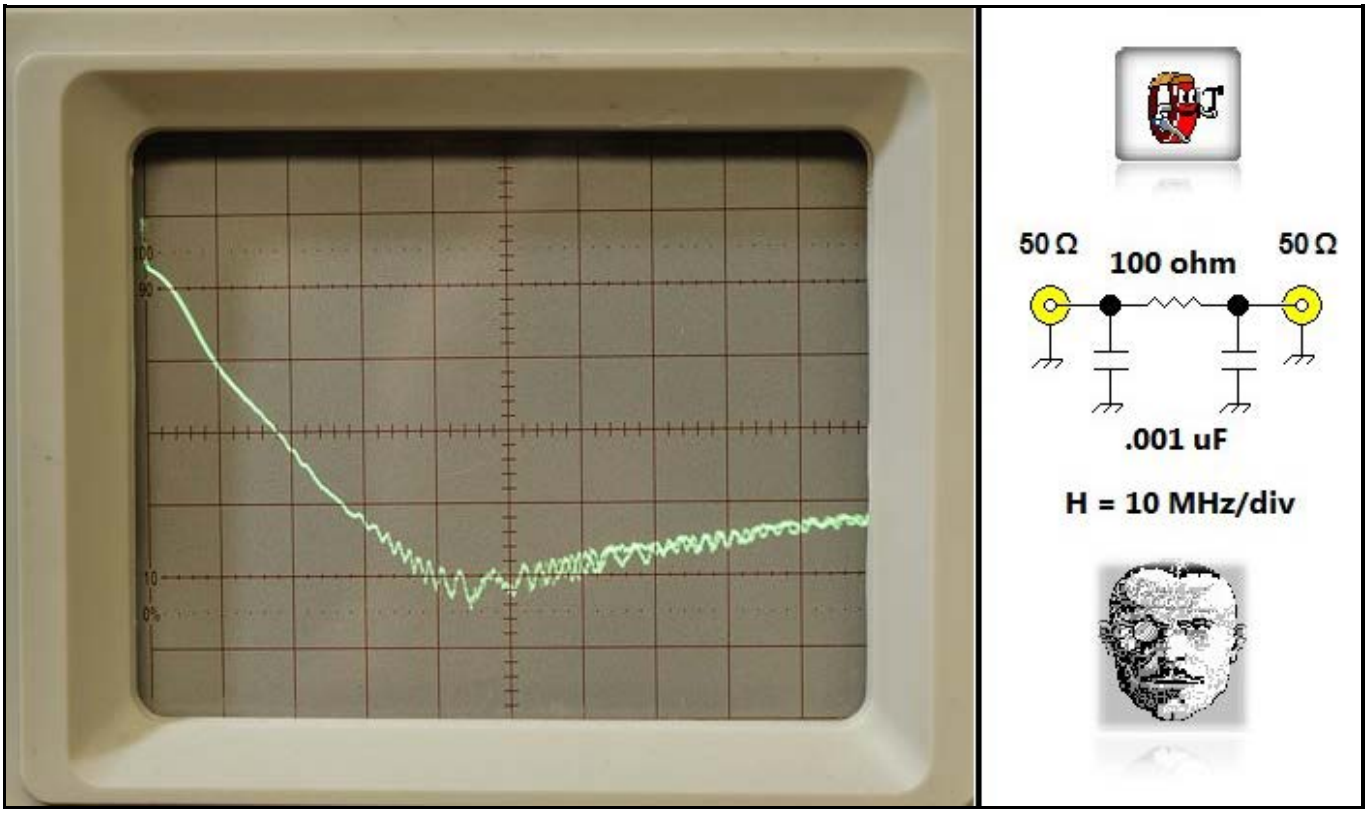
Above — The "bench standard"; a 100 Ω R with a shunt 0.1 µF at each end. We use this all the time. Even at 25 MHz, the attenuation looks around 55 dB down.



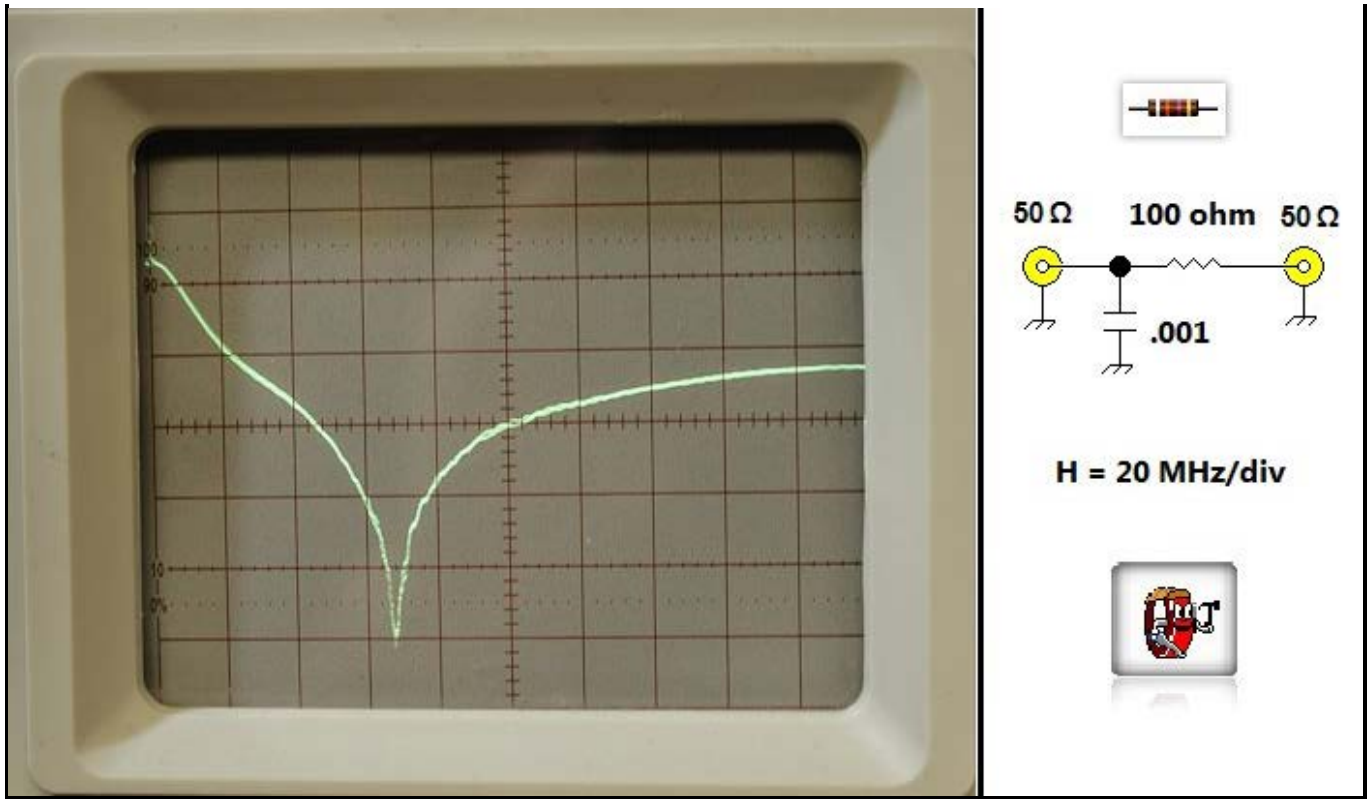
Above — The 100 Ω R with the shunt capacitors decreased to 0.01µF. At 6 MHz, we're about 50 dB down. From 10 to 20 MHz, the attenuation is about as high as I can measure.



Above — 100 Ω R plus 0.01 µF caps out to 50 MHz. I've used this combination of R and C for filtering at 50 MHz a lot.



Above — 100 Ω R plus 0.001 µF capacitors out to 100 MHz. In my particular circuit, the attenuation at 50 MHz equals that of the 100 R + 0.01 µF C low-pass filter shown directly above.



Above — A 100 Ω R plus a single 0.001 µF capacitor. If you leave off 1 capacitor, a serious notch appears at ~ 68 MHz. If you flip the filter around so the bypass cap is on the right hand side, the tracing appears the same. This problem occurred with all the filters tested in all experiments. As possible, solder a suitable bypass capacitor on *both* sides of the R or L.

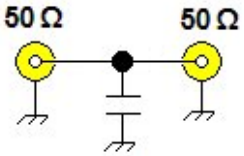
I encourage you to experiment with the SRF of coils and wideband decouple + bypass filters on your own.

QRP Posdata for Oct 2013 — SRF of some common bypass capacitors

Capacitor	Notch Freq
0.1 µF	6.56 MHz
0.01µF	22.68 MHz
0.001 µF	69.62 MHz
0.0022 µF	53.2 MHz
680 pF	87.34 MHz
470 pF	88.17 MHz
220 pF	154.58 MHz
100 pF	222.91 MHz

* Self resonant frequencies of some common bypass capacitors covering ~ 6.6 to 220 MHz

* Use table to inform bypass cap versus frequency choices



Tested as shown to left in calibrated TG + SA using instrument grade 9 cm SMA cables and connectors.

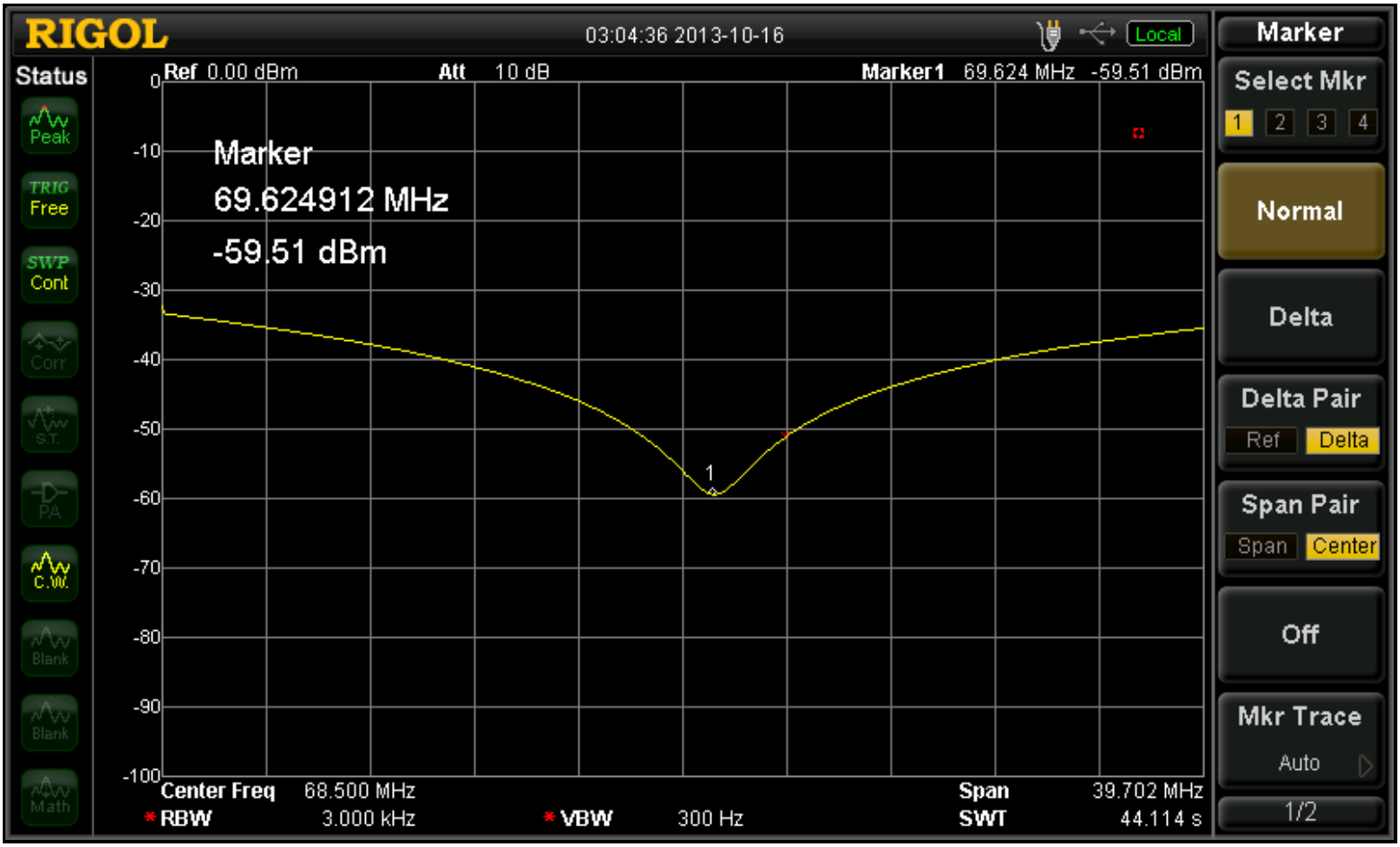
* The deepest notch [-64.32 dBm] occurred with the 0.1 µF cap and decreased progressively with capacitance

* Caps tested = muRata RPE series monolithic ceramic with C0G temp compensation. Different brands, types, etc. of caps may change results significantly

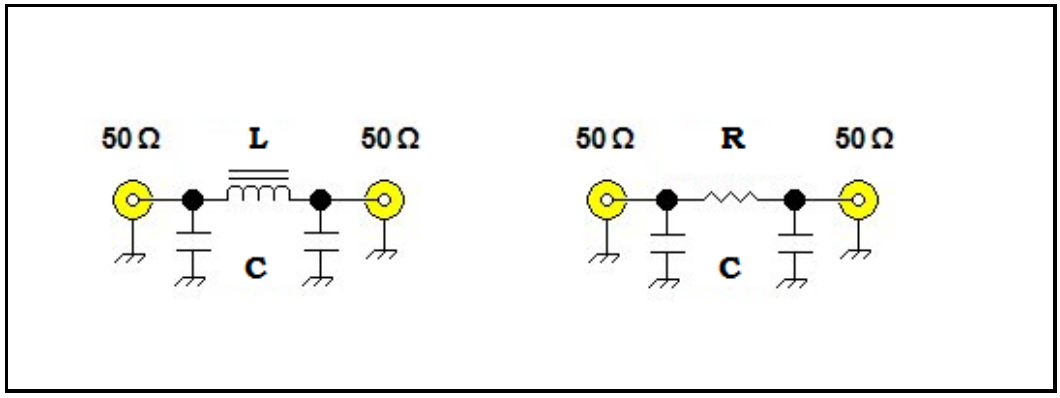
* Leads short as possible with the SMA connectors soldered right on the ground plane

Above — A reference table showing the self resonant frequency of several comon value bypass capacitors in my parts collection. For example, if

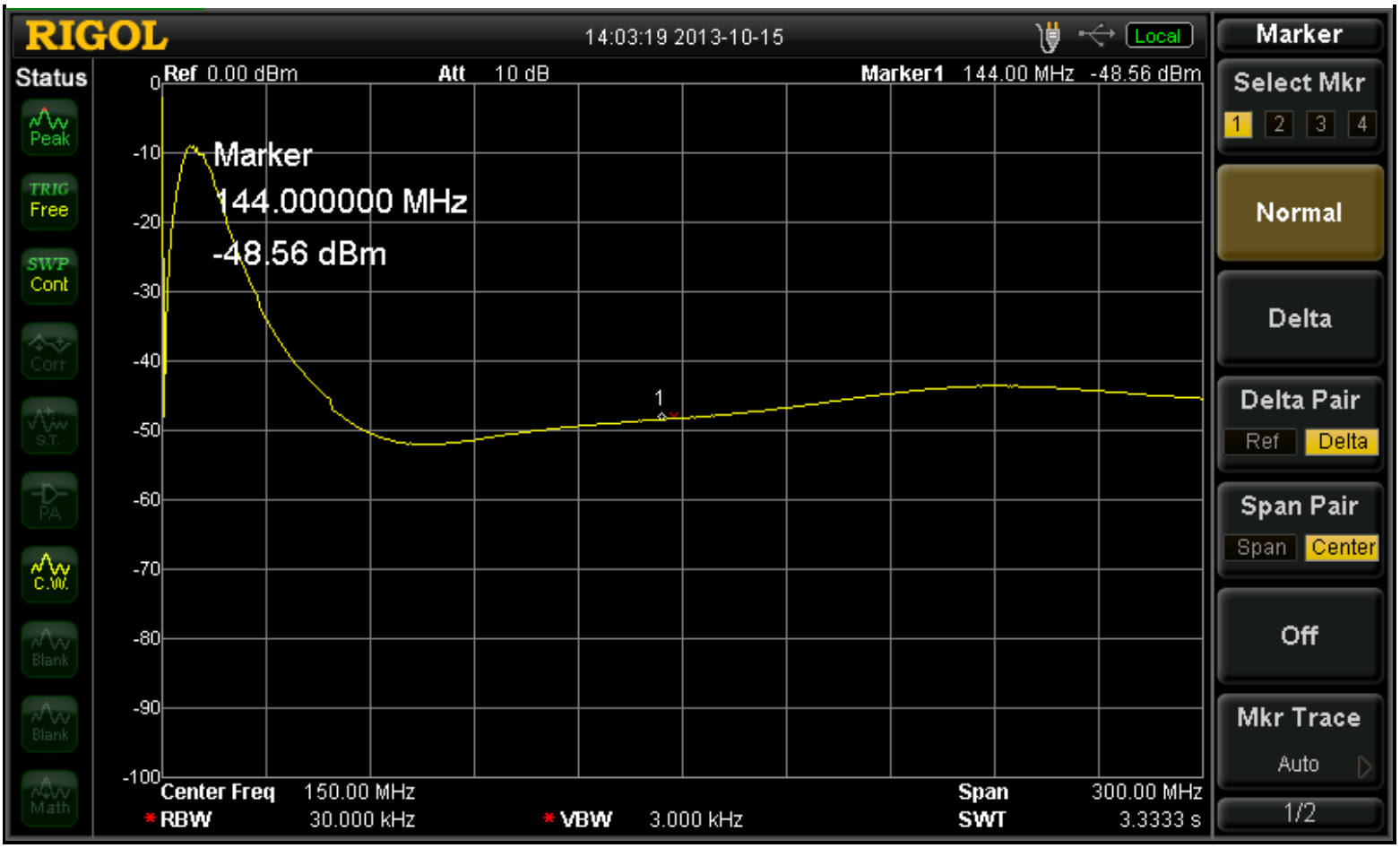
I'm making a 21 MHz circuit, the best bypass capacitor choice from the table above = 0.01 µF. If possible, sweep the capacitors in your own collection to determine their SRF; or whether they're even suitable.



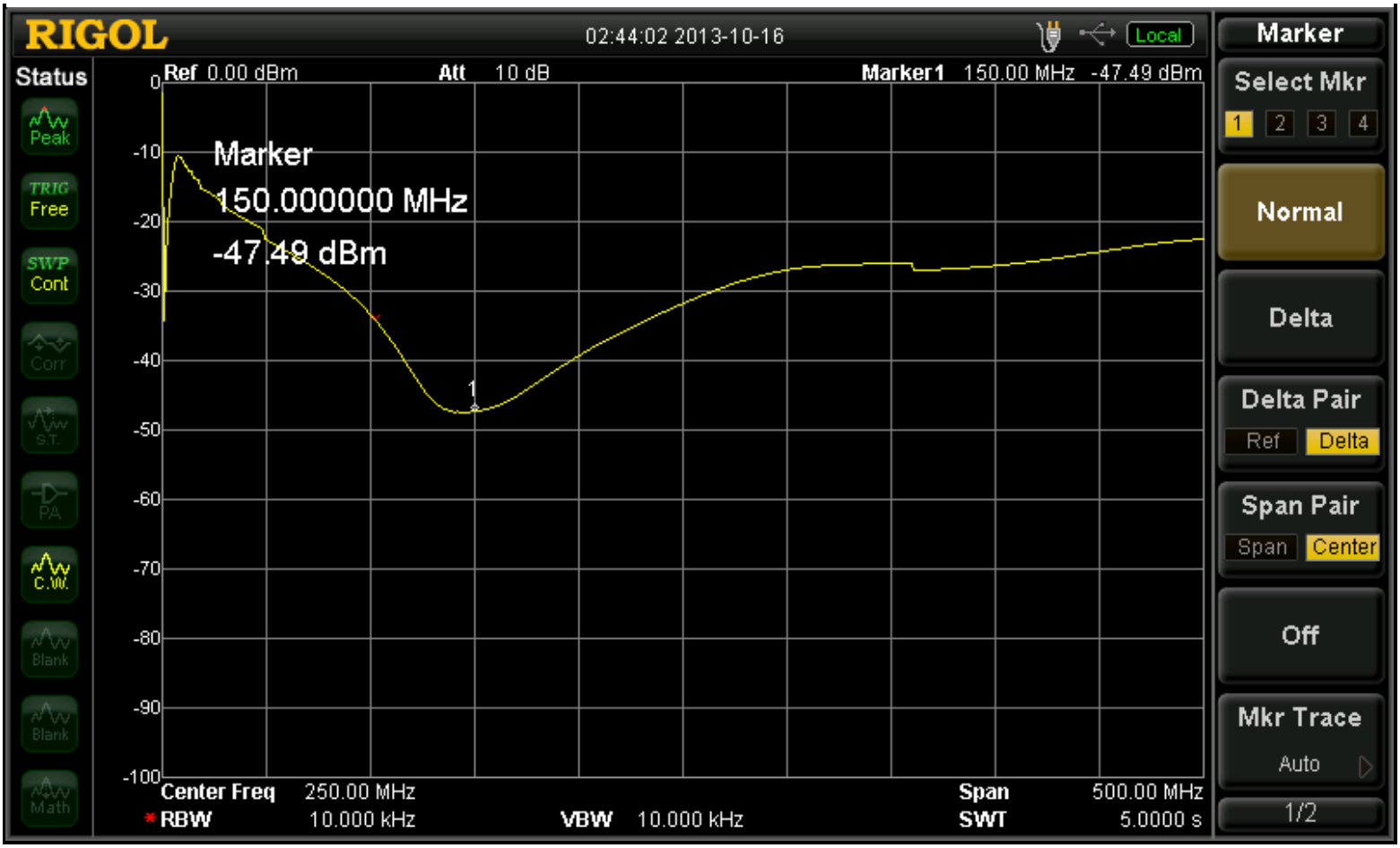
Above — The close-in sweep of the 0.001 µF capacitor tabled above.



Above — As possible, stick 2 of your bypass cap values in a pi filter network with a series decoupling L or R to derive wideband filtration. For example, to filter your DC power lines.



Above — A 300 MHz sweep of a pi filter [220 pF + 1.2 µH + 220 pF] for the DC supply line of a 150 MHz oscillator. The SRF peak lies at 76.69 MHz, but this filter works okay out to about 200 MHz. I placed a marker at 144 MHz and could use this filter for the 2M band as well.



Above — The network described above, except I replaced the 1.2 µH L with a size 0805 10 Ω resistor [220 pF + 10R + 220 pF] and swept to 500 MHz. I set Marker 1 on 150 MHz; the frequency of the oscillator I wanted to DC powerline filter. The resistor gives a bit more filter bandwidth around 150 MHz. A 51 or 100 Ω resistor will further increase the bandwidth while decreasing the attenuation depth somewhat. Although resistors incur a DC voltage drop, they avoid the potential of an unwanted SRF in your filter arising from a renegade inductor— and so, a resistor may pose a better choice for pi filtering DC lines and so forth. It's your call.

At HF and lower VHF, I've found a hole-through capacitor may sometimes filter better than a "garden variety" SMT counterpart. Click for a graphic that shows this. Presumably, the SMT cap exhibited a lower Q than the equivalent hole-though part. At some frequency above 200 MHz, the lead inductance of the hole-through capactor may cause the opposite effect.

Further, on the VHF — Véronique web page, Section 6: I swept 3 capacitors including an ultra-high Q SMT part.

RF Workbench Page 5

Welcome to part 5 of a web series exploring basic RF measurement and bench practices. This installment builds on the information from RF Workbench Parts 1 - 4.

In RFWB #5 I share a hodge-podge of thoughts and circuits concerning power measurement on the beginner-level RF workbench. Consult EMRFD for more support. Big thanks to my mentors: Wes, W7ZOI. Bob, K3NHI and John, K5IRK for their support as I advance to the basics.

Power Measurement Empowers You

Before embracing the 50 Ω RF environment, I misjudged the need to quantify small signal power — now I get that we measure lots of low-level signals on the 50 Ω RF workbench. Whether you're driving a mixer RF port with -30 dBm to reduce spurs, or tweaking an amplifer-under-test to exact the best S21, low-level RF power measurement is fundamental to fruitful RF design.

1 way to measure low-level RF power includes building a log linear RF power meter (PM) based on the Analog Devices AD8307. The basic circuit I show posits that most of you measure from MF to HF and don't need a PM that reads flat into UHF and further; a simple, 2 chip circuit might even prompt you to actually build a barebones PM for your QRP workbench.

Search for and download the Analog Devices AD8307 Revision D datasheet — it's definitely worth a read. Kudos to the design team that brought us a truly milestone device for low-cost power measurement.

Analog Devices offers a whole family of log-amps at different frequency ranges — for example, the AD8311 Log Amp/Detector covers from ~100 MHz to 2500 MHz. A sister product, the AD8302 2.7 GHz RF / IF Gain Phase Detector looks amazing.

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1. A Barebones RF Power Meter

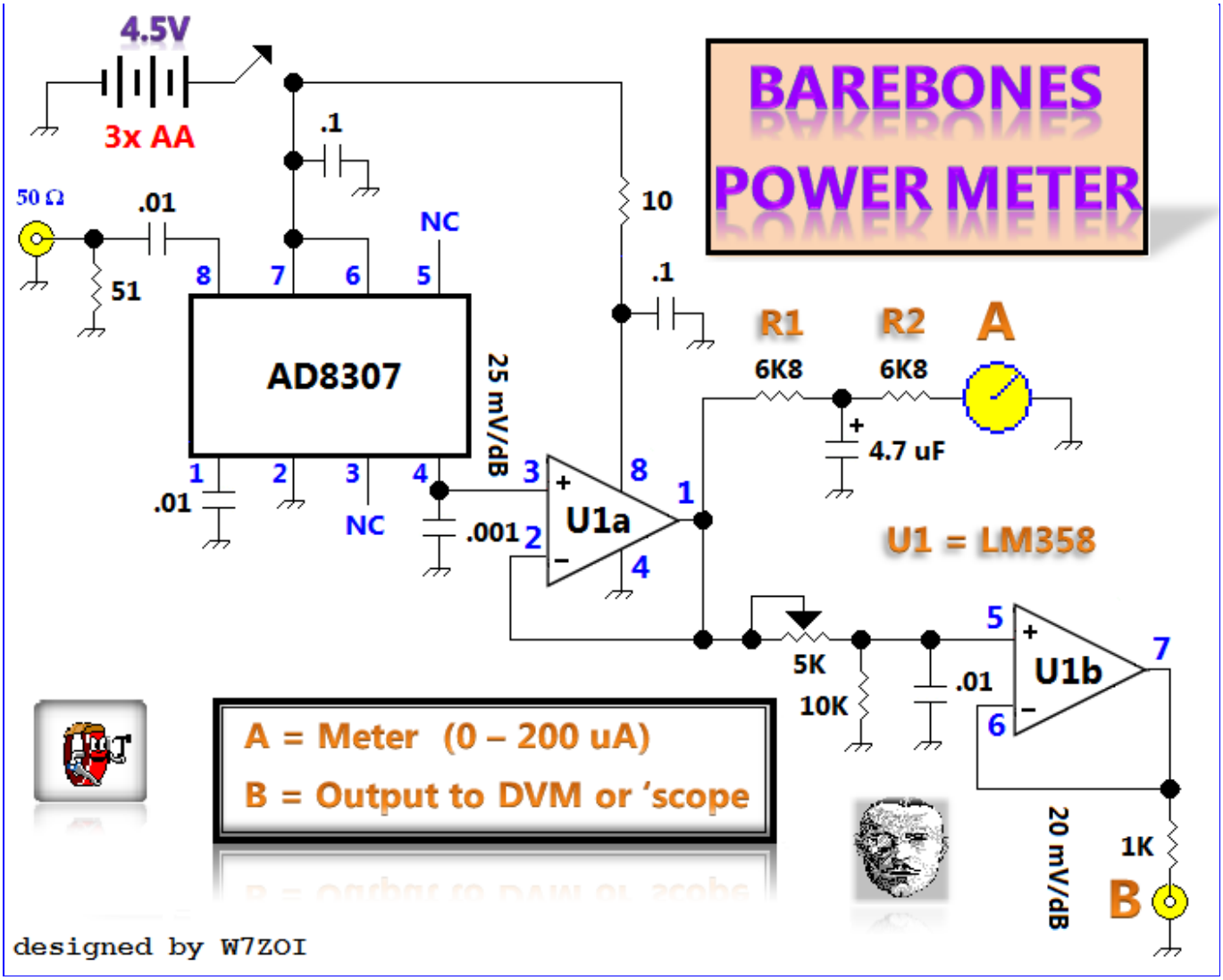
Lacks the usual input frequency compensation network needed to keep power measurement flat over several hundred MHz — since most input compensate by attenuating HF, this simple version offers stronger sensitivity at HF than usual.

Good from MF up to about 100 MHz.

4.5v B+ to eliminate power supply decoupling issues and to give greater battery life than a 9v battery.

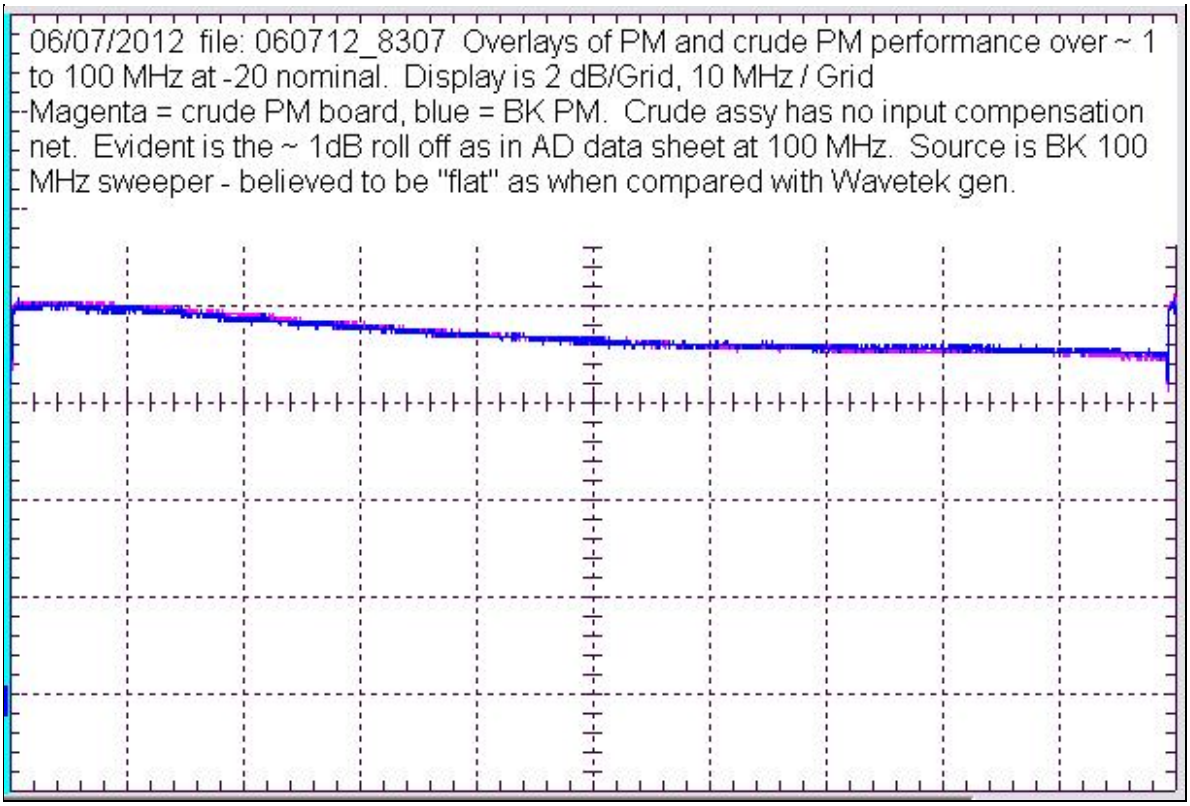
Minimum input power about -70 dBm. Maximum input power +15 dBm. Requires a metal box.

Big thanks to Wes, W7ZOI for letting me present his simple RF power meter. I found *numerous* AD8307-based RF power meter designs in periodicals and on the web and the writers devised simple to elaborate input compensation networks to establish a flat response out to 500 MHz — involving parts such as nH-level inductors, chip resistors and capacitors. This PM goes the opposite direction; a plain circuit for you builders who measure under 100 MHz; particularly at HF.

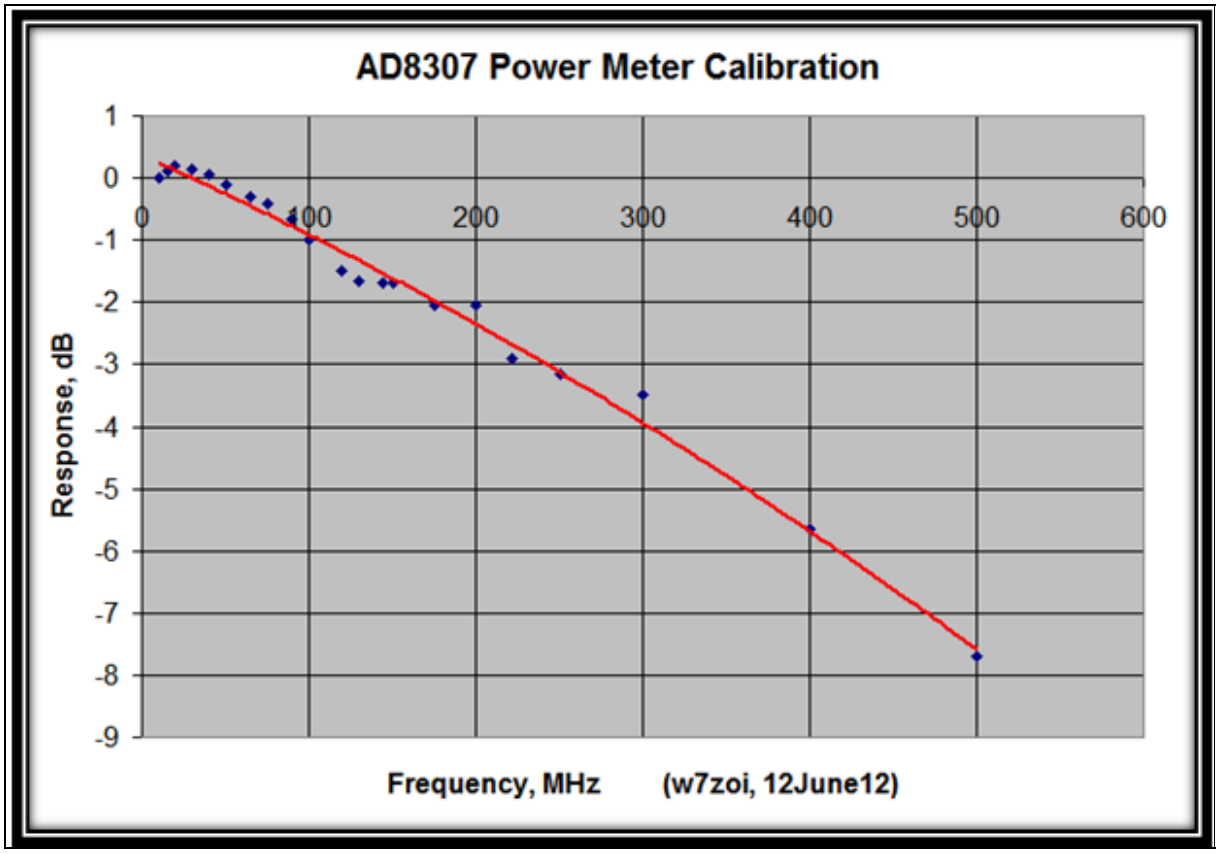


Above — the complete Barebones PM schematic. Presented with the permission of Wes, W7ZOI.

AD8307 Input Pin 8: Some Graphs and Notes:



Above — The lack of input compensation might raise a few eyebrows! A sweep of the Barebones PM by Bob, K3NHI shows a reasonably flat response out to 100 MHz that's comparible to some of the 0-100 MHz range of wider sweeps in the published compensated circuits I've read.



Above — A different Barebones PM plot from Wes, W7ZOI that goes out 500 MHz. For HF work, this power meter proves adequate for QRP HomeBuilders. By all means, add some input compensation if you want to — good circuit examples abound. For example, EMRFD Figure 7.13, or Bob Kopski's — An Advanced VHF Wattmeter referenced in Section 5.

AD8307 specifications allow a typical +/- 0.3 dBm "ripple" in the input to output transfer characteristic. Bob, K3NHI verified this ripple in his lab.

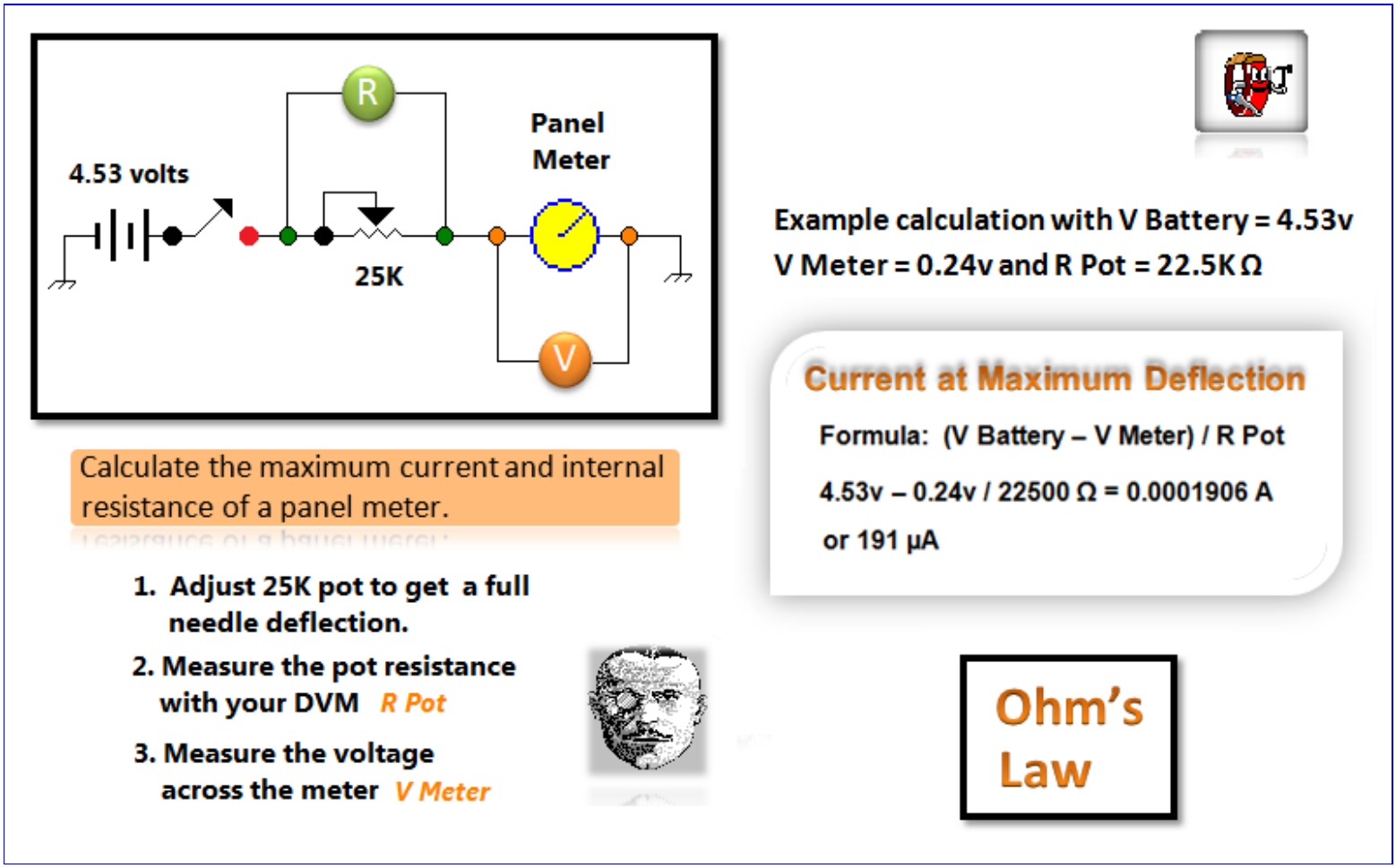
Therefore; depending on the position of a signal in the transfer curve, a move between 2 different power levels may yield as much as 0.6 db peak difference error. Figure 8 of the Rev D datasheet shows this. Figure 24 plus associated text tells why this occurs: the transfer characteristic is a chunk or segmented realization of a log transfer characteristic. It's really good — just not perfect.

If you consider the cost of thermal sensor power meters from Agilent, Techtronix, or even the Mini-Circuits PWR-6GHS, the AD8307 seems a bargain.

Measuring U1a Pin 1 With a Panel Meter

The first output at Pin 1 drives nearly any junkbox DC panel meter you might own The schematic specifies a 0-200 uA movement, however a 1 mA meter movement also works by tweaking R1 and R2 to establish the correct current. In the circuit shown the R = 13.6K. This R establishes the current required to drive the 200 uA meter Wes used in his design — nothing more. A wide variety of panel meters work because the drive current comes from an op-amp. Don't order an expensive panel meter just because you want to employ a 100 or 200 uA model — choose your meter because it affords good needle movement and resolution to allow accurate power measurement.

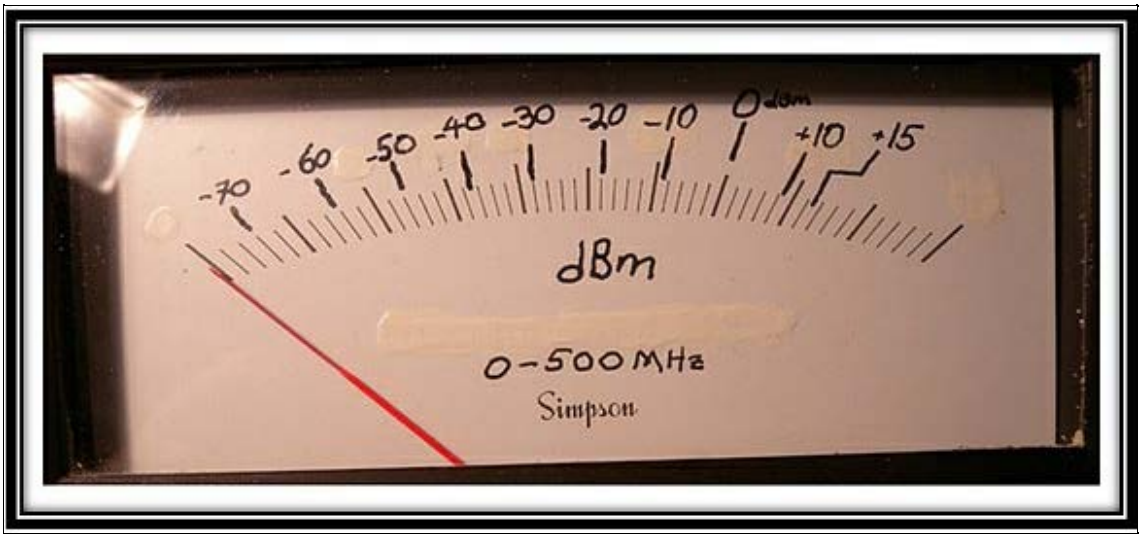
Although elementary for some, the following diagram shows you how to measure the maximum current for your meter: 100 uA, 200 uA, etc.



Above — In my shoebox sat 4 different panel meters. I measured 1 meter and show the math above — a 191 µA meter.

Above — An antique 1 mA panel meter driven with 1 mA to achieve full deflection.

Meter resolution presents the biggest problem for junkbox panel meters and if you look around the web, you'll see some great examples how builders calibrated and/or marked the scale on their AD8307 circuit panel meters.



Above — Panel meter markings on my friend Peter's power meter.

I personally like my panel meters to read about 75% of full deflection at 7 dBm, but it's really your choice. Adjust R1 and/or R2 up or down to give the desired amount of meter deflection for whatever panel meter you own.

You may apply Ohm's law to figure out the maximum in-situ current for any panel meter. In the schematic the total R is 13.6K, so when you apply +10 dBm applied to the power meter input, the current in the panel meter will be 2.0v /13.6K = 147 uA. The maximum 200 uA will occur with a Pin 1 voltage of 2.72v. The op-amp won't go all the way to the positive rail, but If it did go all the way to 4.5v, you would hurl 331 uA into the panel meter. This exceeds the meter's uppermost scale, but likely won't destroy it.

The 4.7 uF cap can be any small uF capacitor value and low-pass filters the DC to smooth out the meter movement. I used a 10 uF in my breadboard.

I normally view the panel meter to tune resonant circuits and observe trends; but not to precisely measure power — it's often more accurate to quantify power readings at Port B since this eliminates panel meter resolution issues.

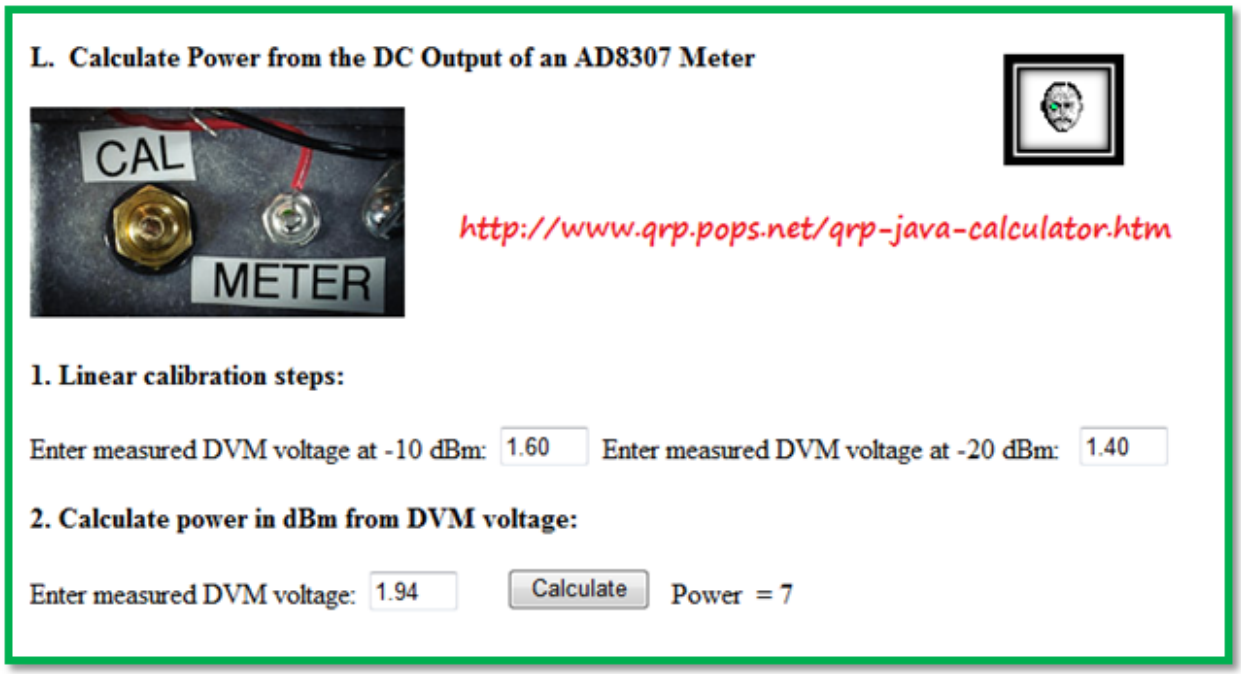
How to Measure Power at Port B

We measure the DC voltage at port B and use equations or graphs to translate this voltage into an actual power reading after reference calibration. U1a and U1b are unity-gain voltage followers to buffer the AD8307 output. U1B features a 5K potentiometer in series with a 10K

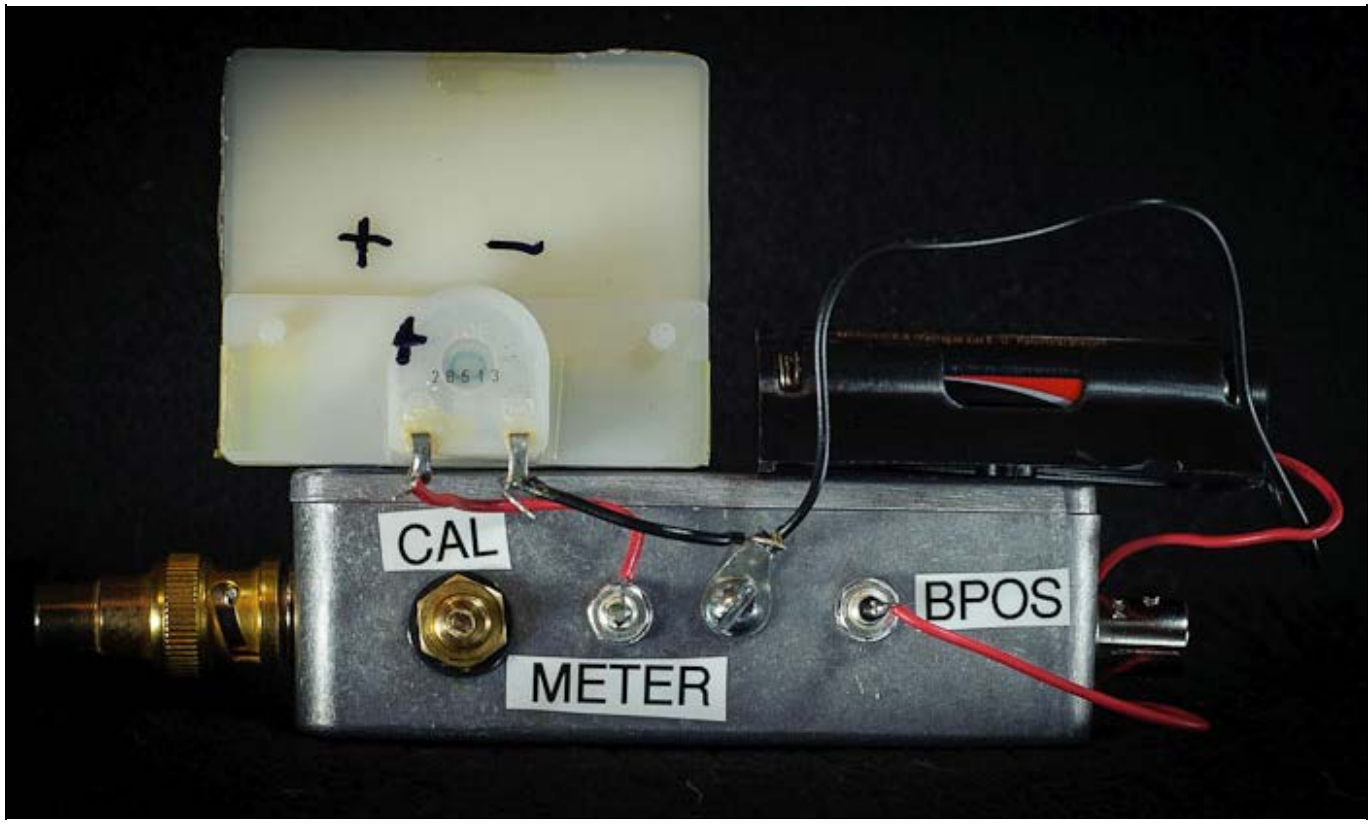
shunt resistor to form a voltage divider that changes the 25 mV/dB AD8307 output to 20 dB/mV. Tweak the 5K pot to calibrate to get as close to 20 dB/mV as possible, although since the LM358 buffers just pass on the DC voltage changes of the AD8307, the pot does not technically alter log linearity.

Wes published some essential notes on his web site. On page 8, you find his formula to convert a measured DC voltage into dBm. I wrote a program that incorporates his formula: Applet L on my Design Center web page; except that it takes calibration power at -10 dBm and -20 dBm.

Let's run through 1 example:



Above — A screen capture of JavaScript Applet L. I calibrated my instrument by connecting a 10.0 MHz sine wave calibrator to the 50 Ω input port of my PM. At -10 dBm input power, my Port B reading was 1.6 VDC; at -20 dBm, I measured 1.4v at Port B. These 2 points establish the log linearity per Wes' notes. After this calibration, I measured the output power of a 4 MHz sine wave VFO that I designed to run at 7 dBm output power to drive the LO port of a diode ring mixer. Port B read 1.94 volts and when I entered this into Applet L and pushed <<Calculate>>, the VFO output power was indeed 7 dBm. More on this later...



Above — My prototype RF-tight breadboard of the Barebones PM. I ran my 'trademark' die case box, BNC RF input and Port B connectors, a common ground lug and feedthrough capacitors for the B+ and panel meter connections. I placed a BNC to RCA adaptor on Port B to allow the insertion of a standard-type positive DVM probe. A black alligator clip terminates the distal end of the negative probe on all my DVMs. I just clip it onto the ground lug.

Above — A view of my breadboard showing my 100 uA panel meter and the 4.5v battery pack. I raised R2 in the original schematic up to 8K2 to set my preferred needle movement in this 100 uA panel meter. A meter with linear markings might be better?

Some Further AD8307 Notes

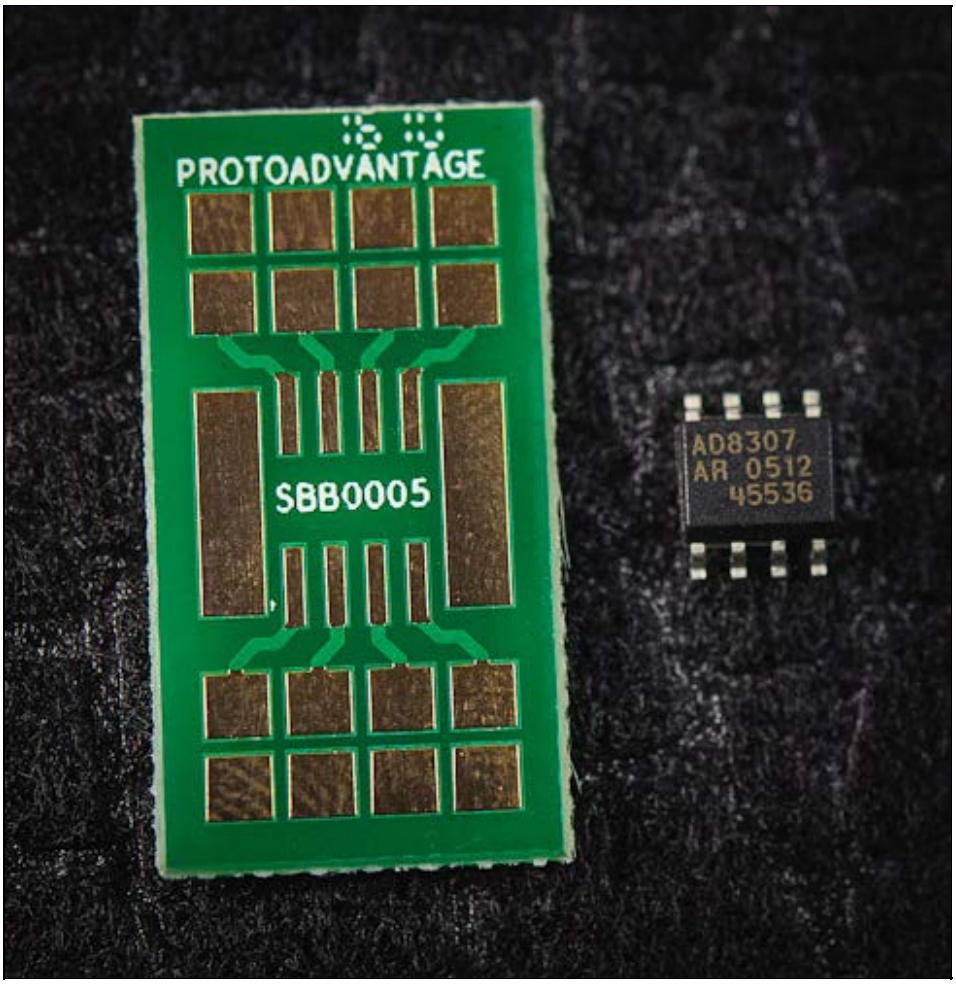
After power up, you'll notice a DC output voltage around 0.22v or so with no input signal; this arises from wideband noise caused by resistors and

amplifiers in the AD8307— all normal.

If possible, verify a 25 mV per dB power change with a manual sweep using a sine wave signal generator on the input and a DVM to AD8307 Pin 4. If you don't have the test gear to perform this function check, no problem.

The AD8307 input resistance = 1100 Ω , so we must place a resistor in parallel with the input to establish 50 Ω. Many builders just shunt a short-leaded 51 Ω resistor from input to ground like Wes did, however, you might also see builders place a 1% tolerance 53.2 Ω R in that slot to derive an input Z of 49.9 Ω. I did this in another "blinged-out", frequency compensated AD8307 meter I built for future UHF circuit experiments.

With 20 and/or 40 dB taps, attenuator pads and some 50 Ω cables, a Barebones AD8307 PM can measure everything HF you might build or buy for your *QRP* workbench. I posit that a simple AD8307 power meter may form the heart of a basic, first QRP workbench. Lacking a oscilloscope when I started in radio electronics, I measured RF with a germanium diode RF probe and a DVM — I would have enjoyed a simple log power meter plus a basic calibrator, however, the AD8307 did not exist back then.



Above — Some builders lament because an AD8307 (in DIP) costs around $10.00. I bid for and bought the AR version chip shown above for $4.42 USD. The above green Proto Advantage breakout board cost $1.00. To compare; some people spend $5.00 for a boutique coffee in Canada — it's all good.

The SMD packaged AD8307AR may offer better performance above 100 MHz with its lower lead inductance.

QRP — PosData for Dec 3, 2013

Realize your bench acumen — adapt and build some test equipment to suit *your* needs. Steve, VK2SJA crafted a version of the AD8307 PM. Click for his beautiful box featuring a 1mA meter movement. Aided by the Tonne software "Meter" program and the AD8307 datasheet, his ranks among the best I've seen. Kudos Steve.

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2. Power Meter Calibrators

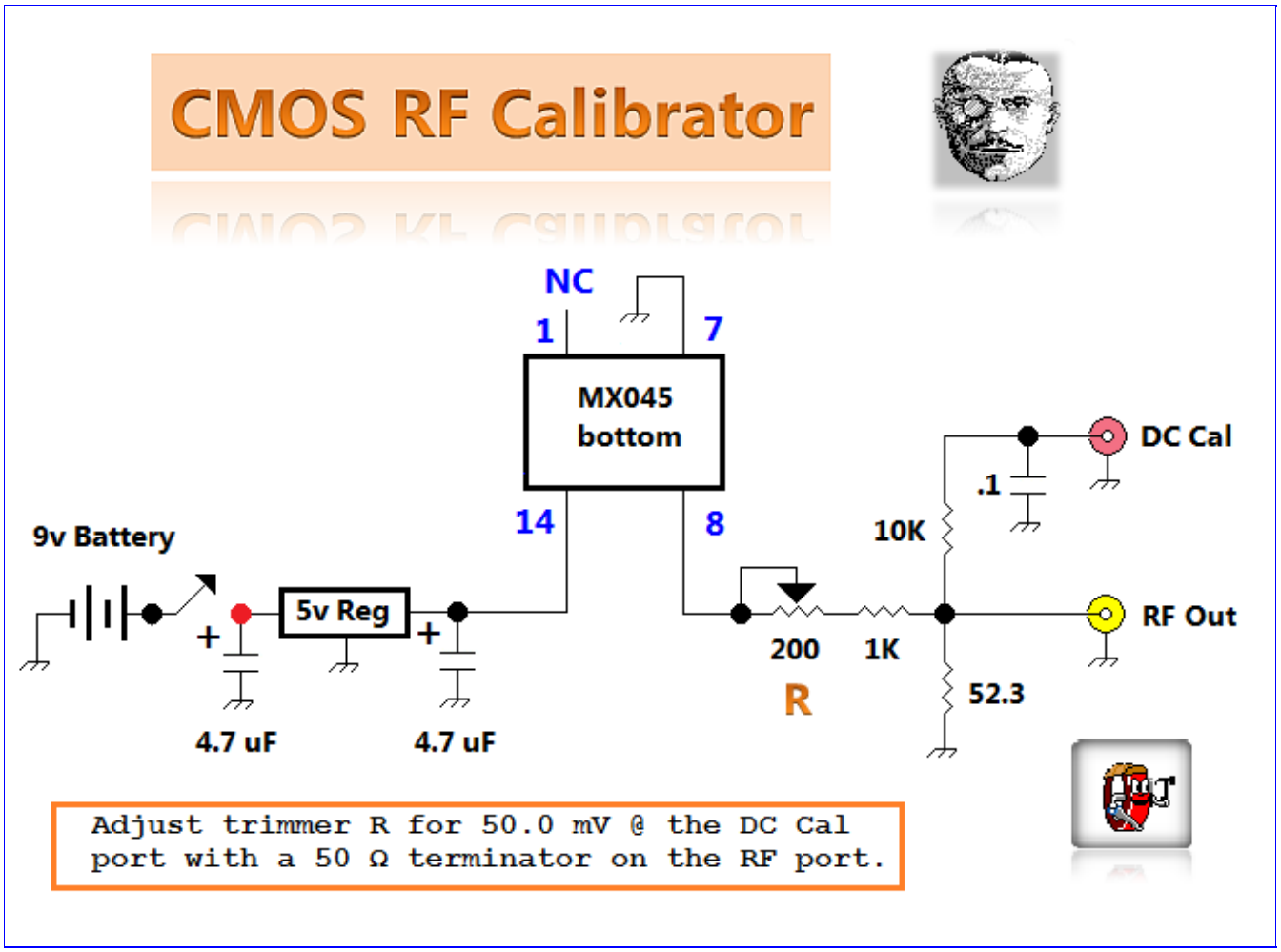
1. CMOS Clock Oscillator RF Calibrator @ -10 dBm

Bob, K3NHI designed a 10.0 MHz reference oscillator for -10 dBm that might help equip the beginner bench. This CMOS signal source does not

need an AC power reference to calibrate it — just a DVM to make a DC measurement and your good to go!

In addition to calibrating an AD8307-based PM, you might use this -10 dBm reference to calibrate other gear including homebrew sine wave signal generators; that's what I'll do later.

Above — The recommended CMOS clock oscillator for the CMOS RF calibrator. Digi-Key part # CTX772-ND. Data sheet.
If you substitute another CMOS clock, it must swing nearly rail-to-rail for accuracy. Bob intended this calibrator for those who lack the bench instruments needed to precisely calibrate RF devices since only a DVM is needed for calibration. This generator also offers a range of calibrated harmonics as Bob described in his January-February QEX article A Simple RF Power Calibrator — great when you want to examine a spectrum analyzer over a limited span.



Above — My version of the K3NHI CMOS signal source (presented with the permission of Bob, K3NHI). The 52.3 Ω resistor is a standard 1% part and I bought 5 for all my AD8307 projects and this little signal generator. In my first version, the trimmer resistor was 500 Ω and worked okay, but the 200 Ω trimmer improved calibration. I ordered 10 Bourns 200 Ω and 10 Bourns 500 Ω trimmers on eBay for a few dollars and after studying Bob's published work and applying his influences to my own, I now love to precisely calibrate or bias circuits with a 200 or 500 Ω trimmer R as appropriate.

QRP — PosData for November 22, 2012

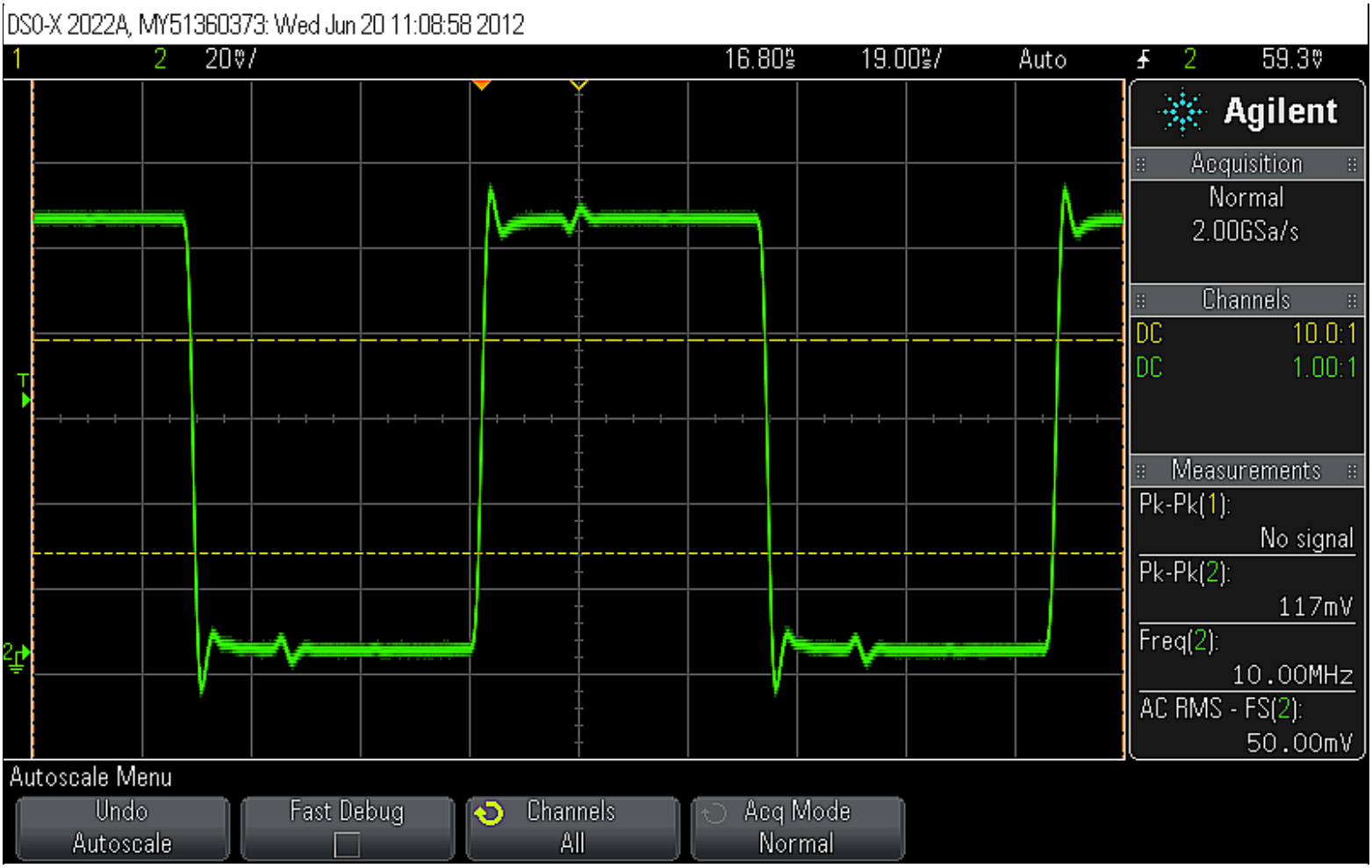
If required, you may substitute a 49.9 to 51 Ω resistor for the 52.3 Ω specified.

Per Bob's QEX 2010 Tech Notes and his emails specifically about calibrating the AD8307 with a CMOS *square wave*: Modern AD8307 chips are better calibrated with a -10 dBM square wave. To test log linearity after CMOS signal generator calibration, apply a sine wave signal generator to your power meter and adjust its output to get the same power meter DC output voltage as with the CMOS generator. Then insert attenuator pads on the now calibrated sine wave generator to assess the mV/dB change with different power levels.

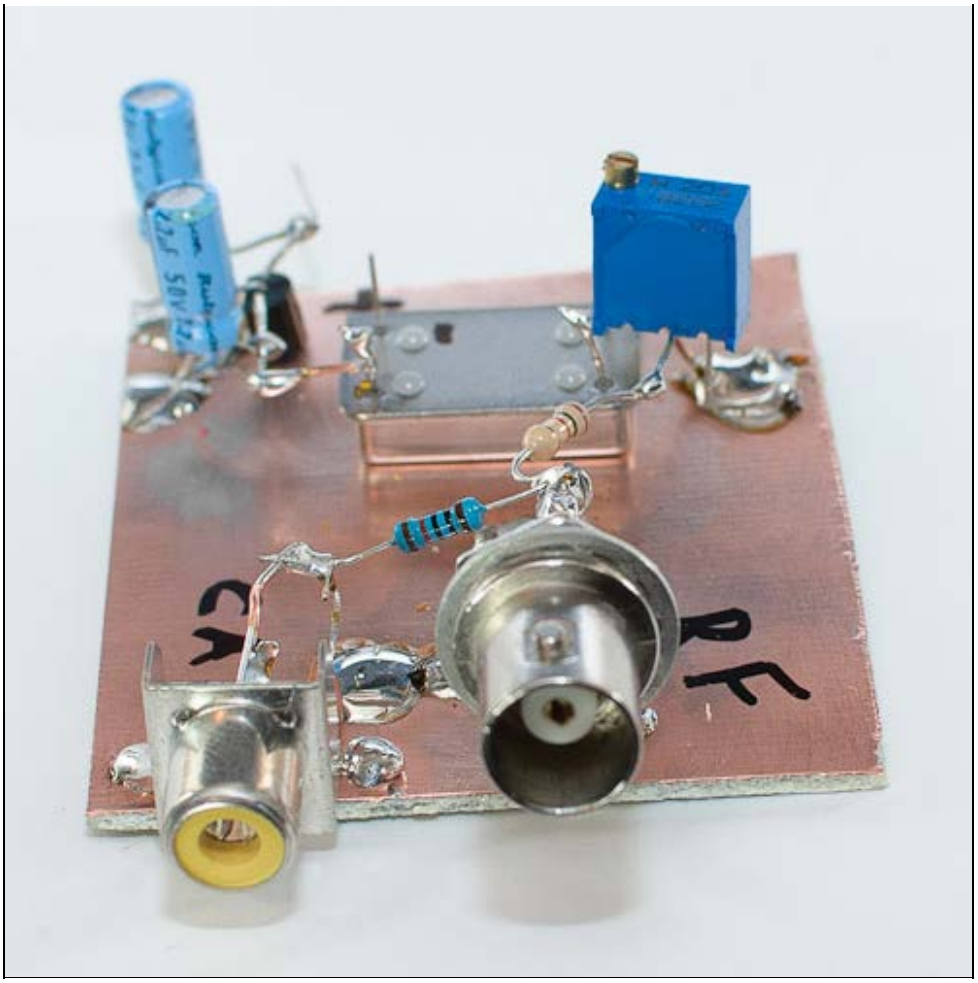
Bob's original CMOS calibrator outputted - 20 dBm, however, he updated it to output -10 dBm in 2010 as reported in his QEX 2010 Tech Notes.

Some readers have asked why calibrate the AD8307 at 10 MHz? Calibrate at whatever frequency you want, or more than one. However, at 10 MHz, the AD8307 exhibits its best log performance compared to other frequencies. Click or click for datasheet graphs.

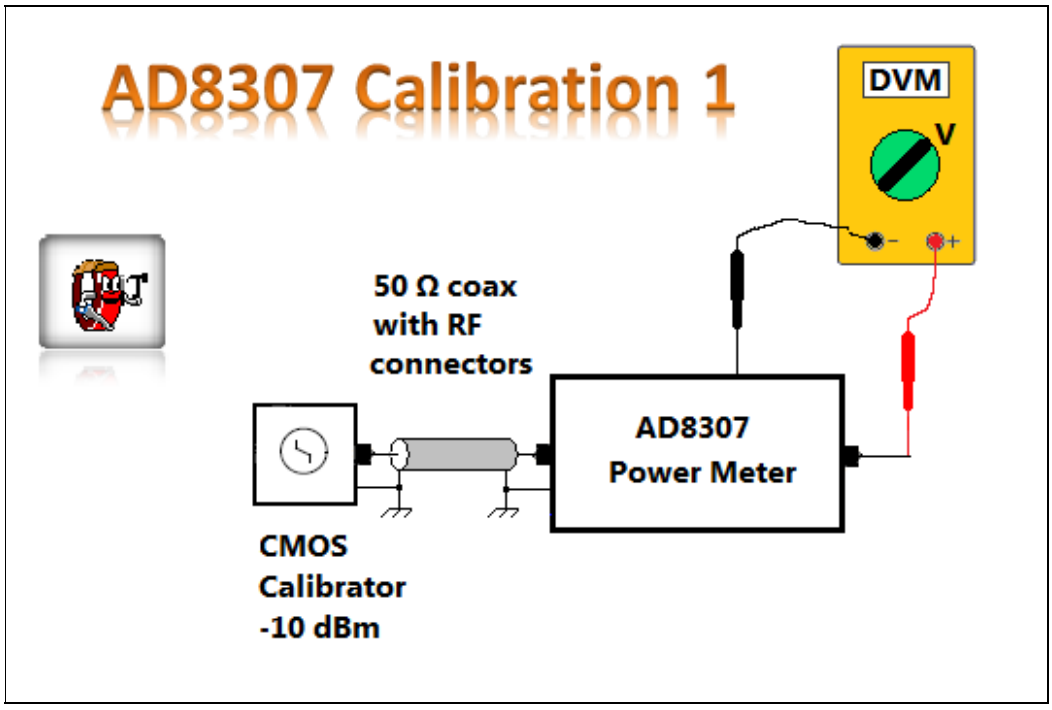
Above — The DVM calibration port reading (2 volt scale) from my CMOS clock calibrator after calibration. That was easy!



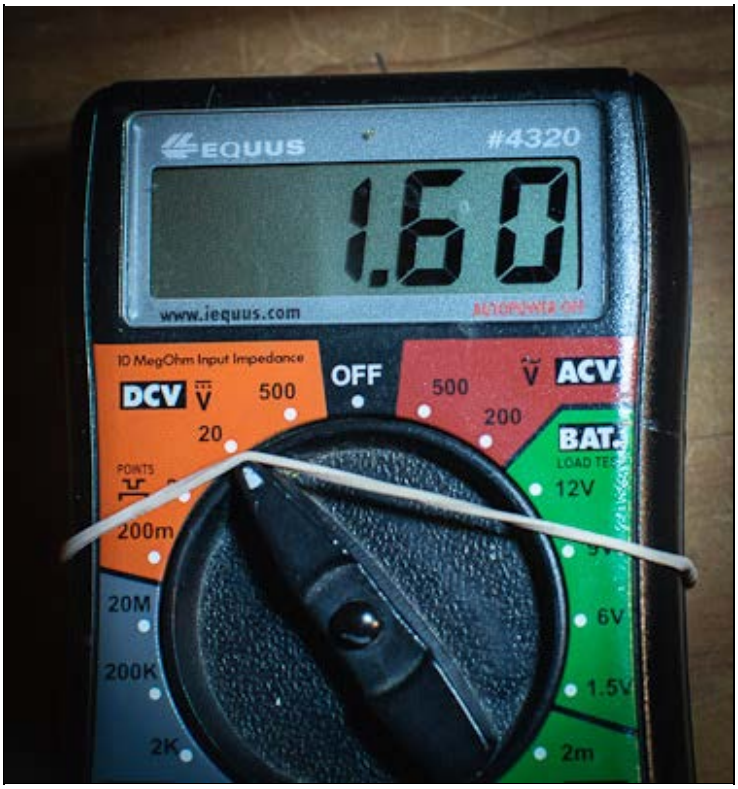
Above — The calibrated CMOS clock RF calibrator in my 50 Ω terminated scope.



Above — The breadboard of my version of the K3HNI CMOS -10 dBm RF calibrator. For best results, stick it in a shielded box. When correctly calibrated, it outputs -10 dBm on an AD8307 PM, -14 dBm on a spectrum analyzer and -13 dBm on a conventional, thermal-sensor power meter.



Above — The procedure for calibrating an AD8307 PM with the -10 dBm square wave CMOS signal source.



Above — When I connected the calibrated CMOS clock RF calibrator to my build of the Barebones PM, I measured 1.60v with my DVM; ***the -10 dBm reference voltage.*** Since the log-linear power changes 20 mV/dB; for my calibration reference voltage: 1.40v = -20 dBm, 1.20v = -30 dBm, 2.0v = 10 dBm etc.

Remember that the Barebones PM runs on a battery pack and over time the B+ will change. *Each time* I measure power with the meter I first calibrate it to establish the -10 dBm reference voltage.

Here's a simple formula that only works for 20 mV/dB @ my particular 1.60v calibration voltage, but gives you the general idea:

Power in dBm = 50 x (V - 1.8)

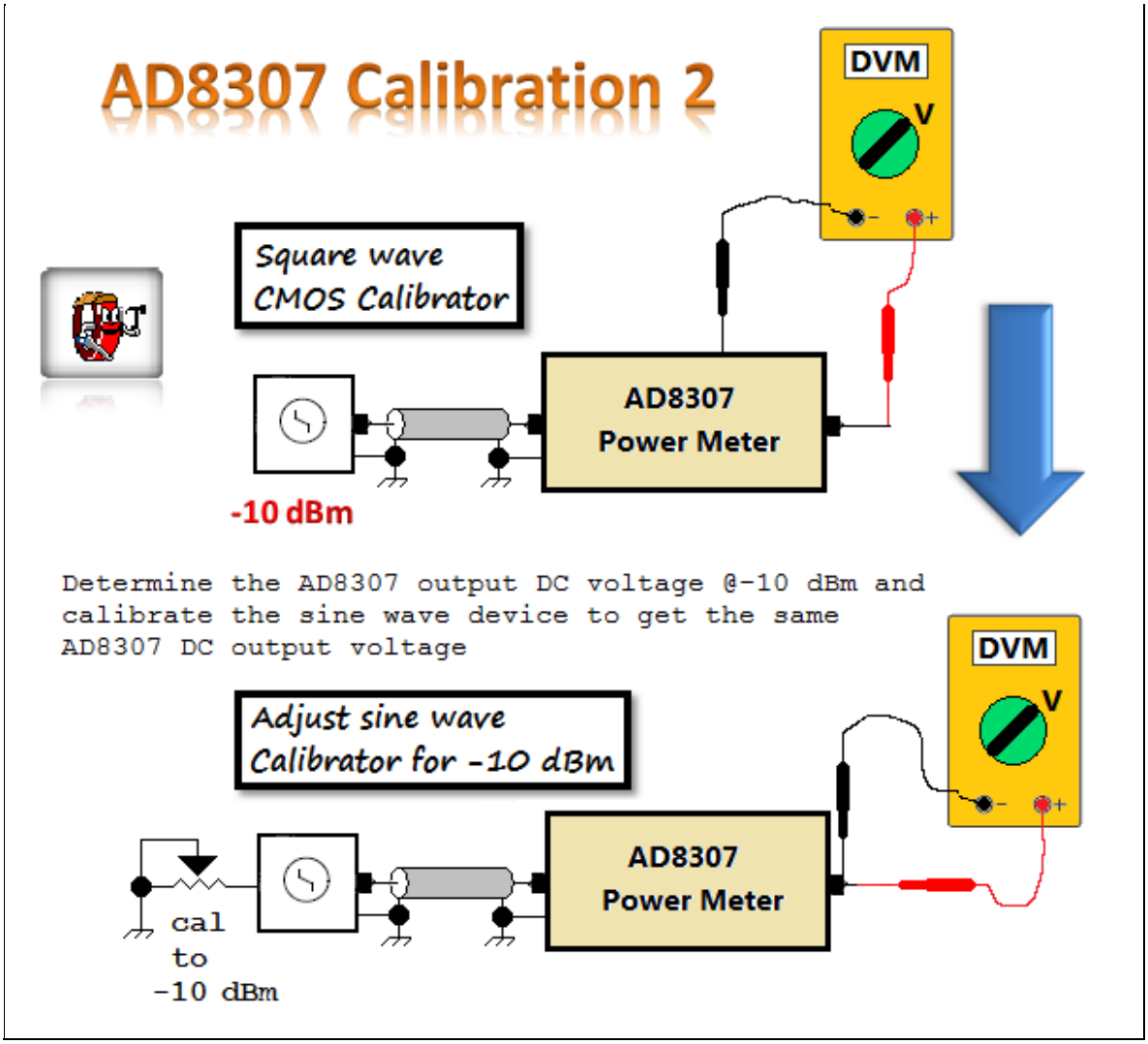
So if I measure 1.94v: 50 x (1.94 - 1.8) = 7 dBm.

2. Sine Wave Oscillator for Calibration @ -10 dBm and -20 dBm.

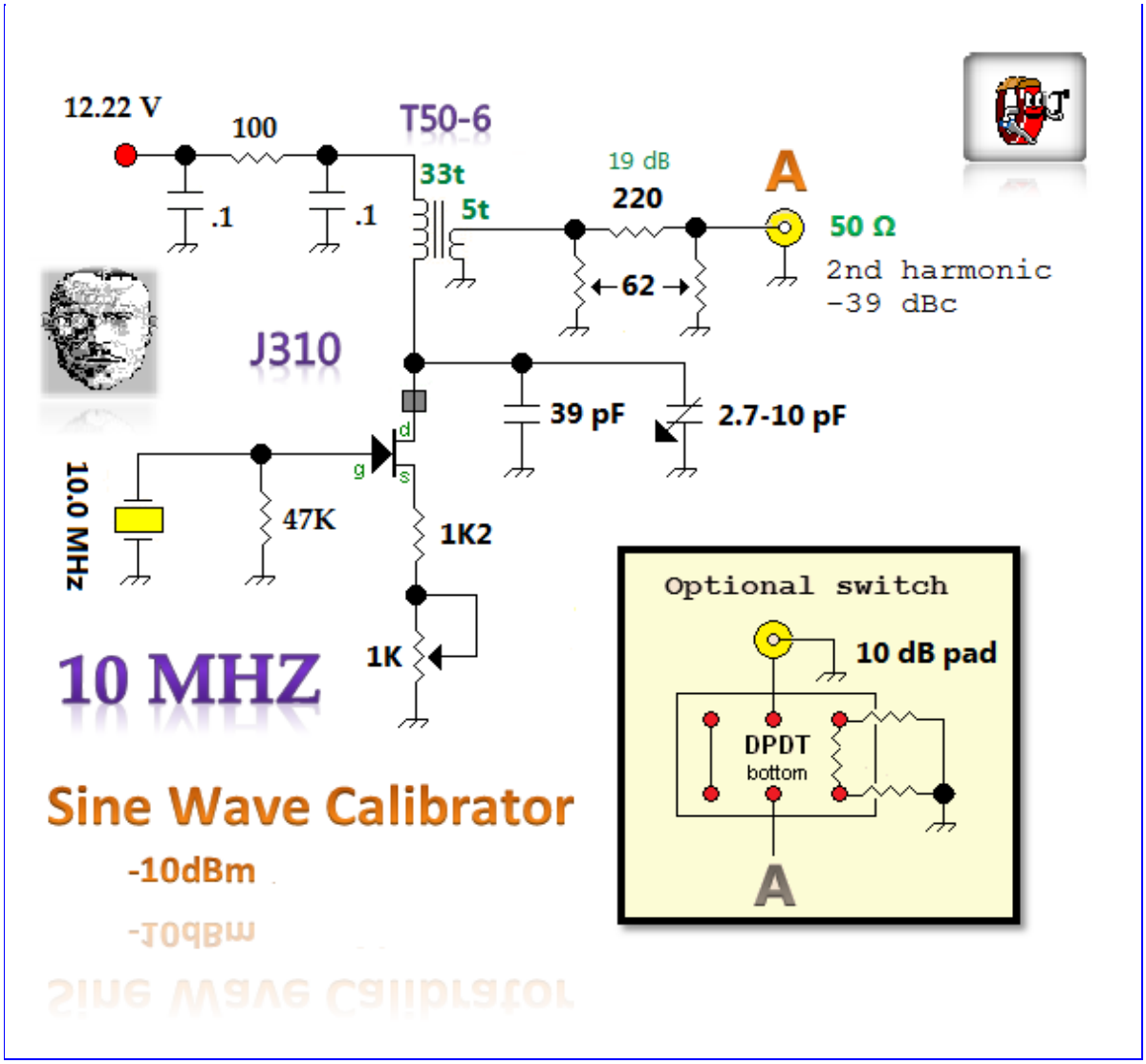
You may also calibrate your AD8307 PM with a calibrated sine wave signal generator. Advanced builders who own the gear needed to measure RF power tend to use a sine wave for calibration.

It's easy to calibrate the AD8307 with a sine wave signal source. Normally we calibrate our sine wave signal generators with instruments such as a 50 Ω terminated 'scope, a spectrum analyzer, a calibrated power meter, or a 49.9 to 51 Ω terminating resistor plus a 10X probe etc., but if you lack these instruments, your stuck.

No problem. You may calibrate any appropriate sine wave oscillator at -10 dBm with your Barebones PM and the CMOS RF Calibrator shown earlier. Let's examine the procedure:



Calibrate the AD8307 PM with the CMOS square wave reference and record the DC voltage at the output of Port B. Connect up your sine wave generator and adjust its output until you get the same Port B DC voltage as the reference CMOS RF signal generator — your sine wave oscillator should now be calibrated to -10 dBm.

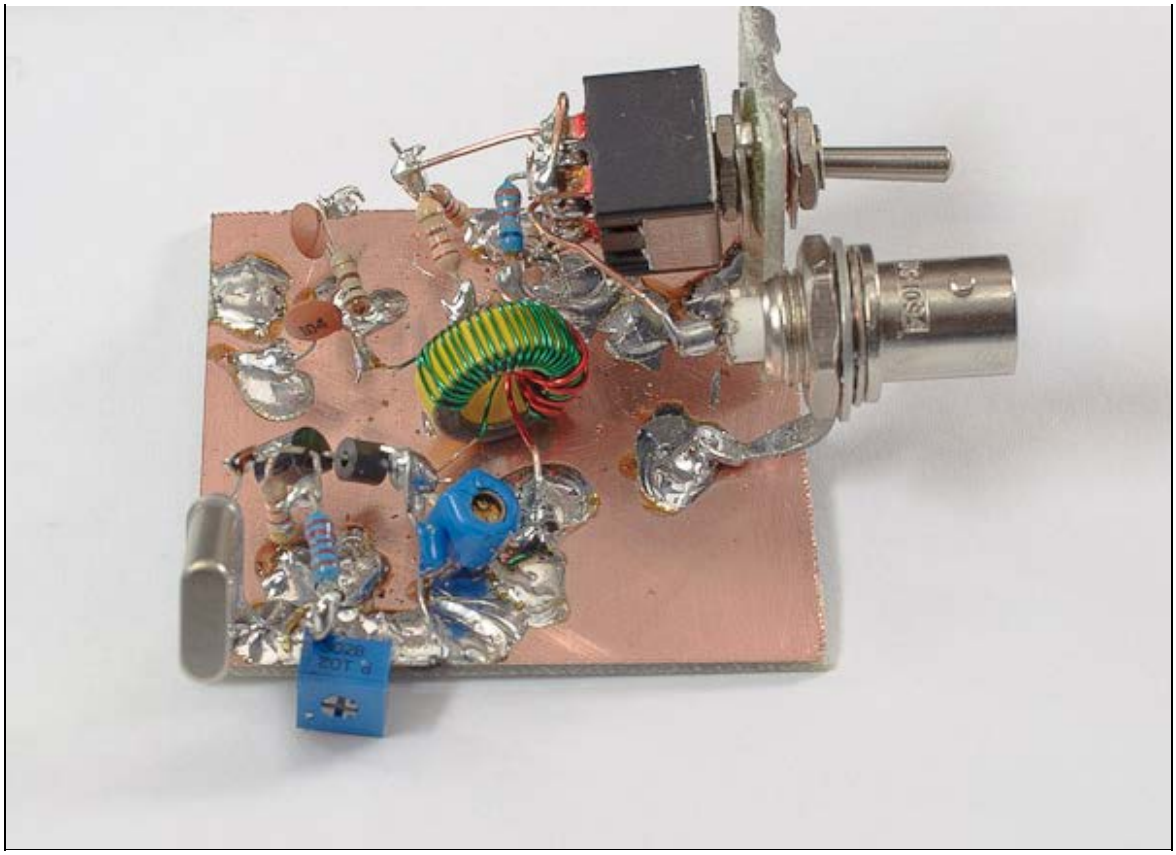


Above — I designed this simple 10.0 MHz sine wave calibrator to evaluate Bob's CMOS square wave RF calibrator and serve as an example sine wave reference oscillator. When running a regulated VCC of at least 12v and the *tank is perfectly* tuned, low distortion arises. I measured the second harmonic @ 39 dBc down. The L with 33 turns = 4.43 uH.

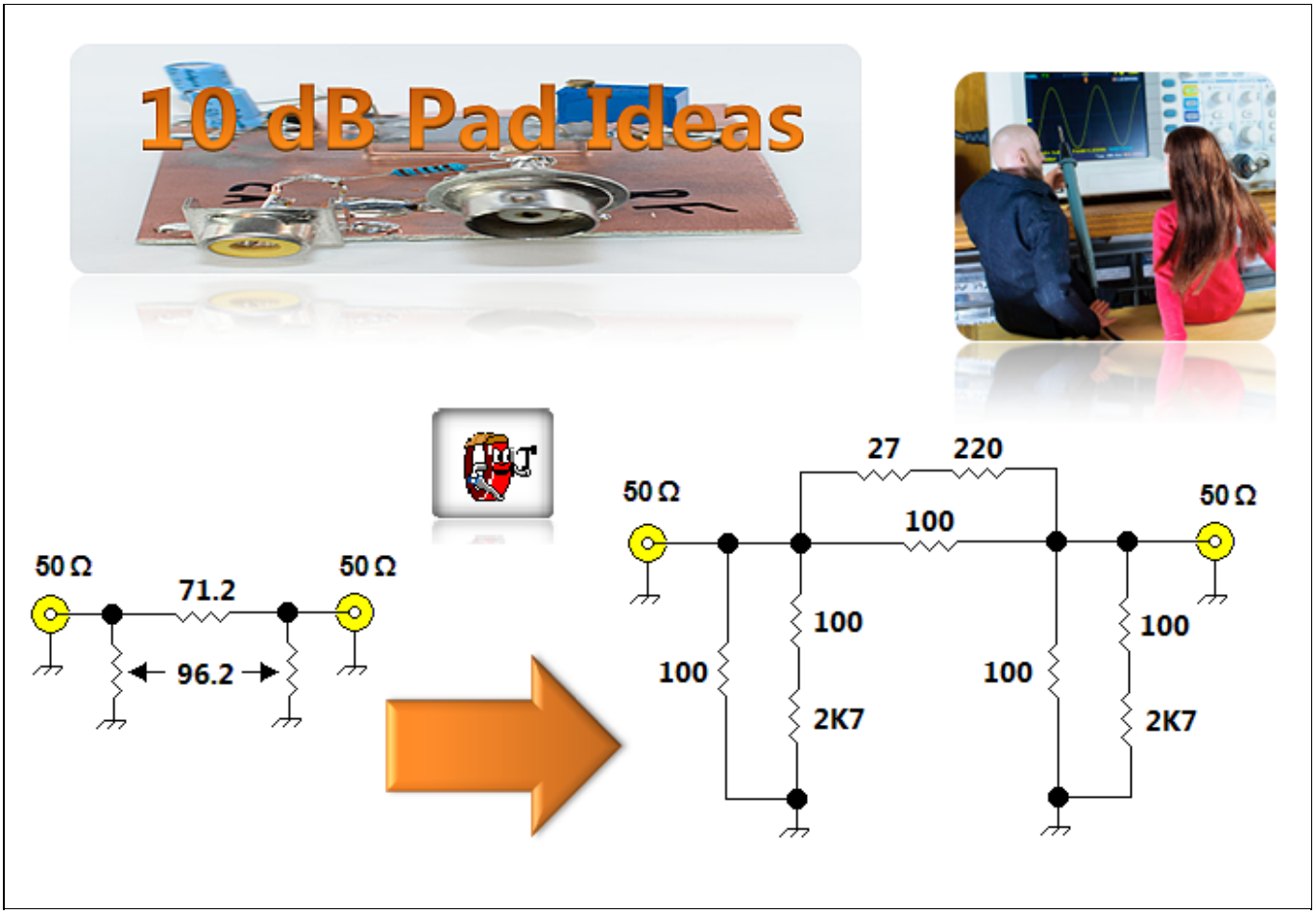
The initial tuning procedure goes like this:

1. Terminate the output with a 50 Ω resistor terminator, or a 50 Ω terminated 'scope, or your AD8307 PM.
2. Connect an ammeter between the VCC node and the regulated power supply.
3. Adjust the emitter trimmer R so that the circuit draws around 2.7 mA — then disconnect the ammeter leads.
4. Adjust the trimmer cap for the highest pk-pk voltage (and/or or best looking waveform) in the 'scope, or highest power in the AD8307 PM. *Nominal total C to resonate my particular circuit was ~ 45 pF.*

If not already done, connect the tuned-up sine wave signal source to a AD8307 PM. Adjust the trimmer potentiometer so the Port B DC voltage = the reference voltage measured during your CMOS RF generator power meter calibration. Your sine wave signal source is now calibrated to -10 dBm.

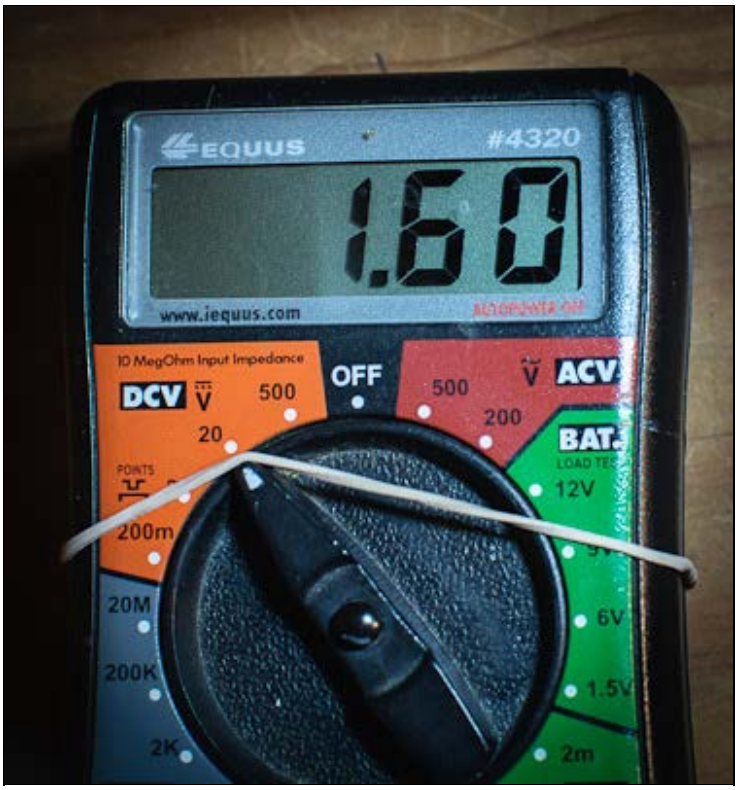


Above — The breadboard of my 10 MHz sine wave signal source. I included the optional switched 10 dB attenuator shown on the schematic inset. In the end I decided to just stick a removable **10 dB** pad [like this 6 dB pad] in-line for my -20 dBm measurement.

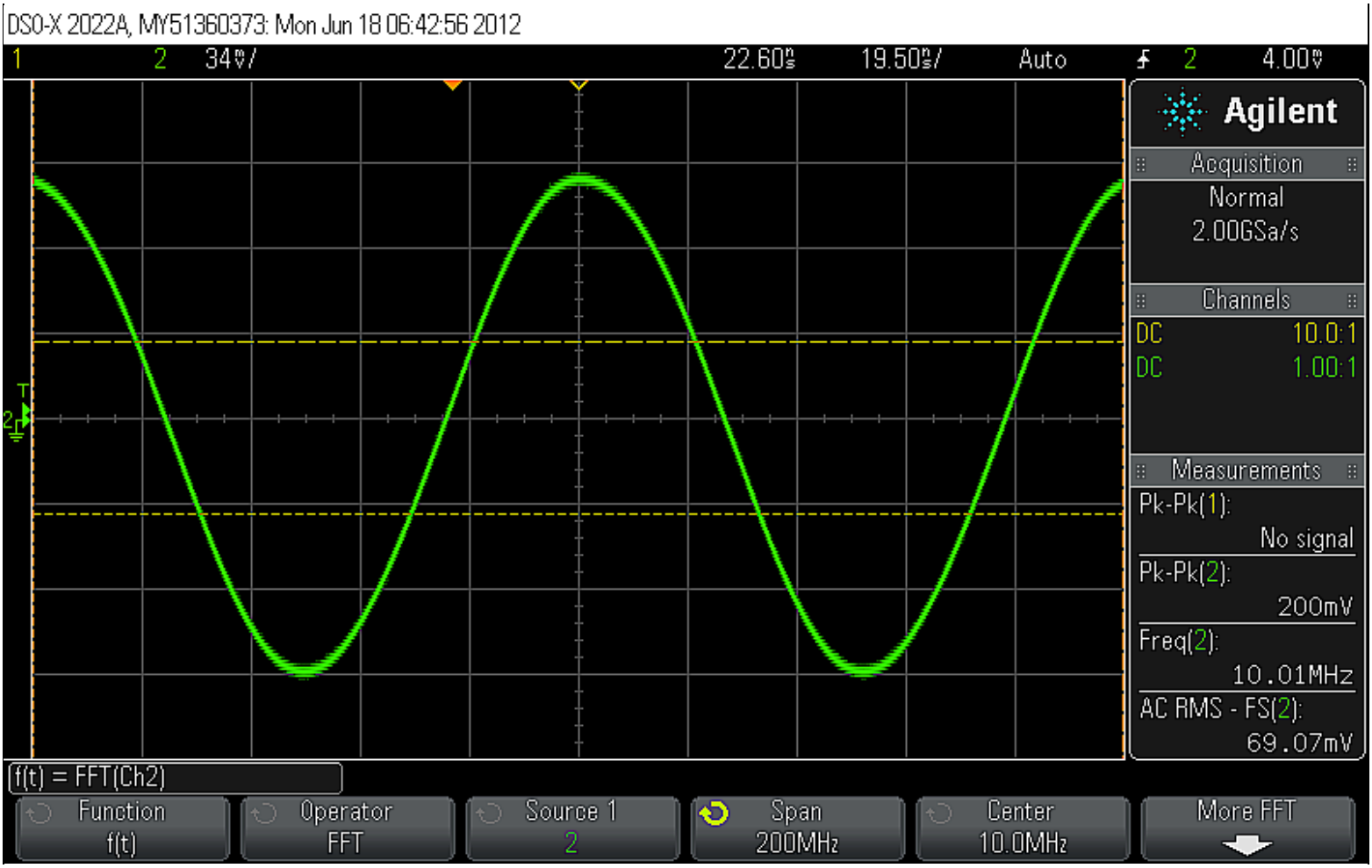


Above — Craft accurate attenuator pads with parallel + series resistors and some creative energy. Lacking the proper 1% parts, I hand selected some 5% resistors among the values shown to the right and built a pad that *precisely* gave a 10 dB power drop at 50 Ω.

Now let's check the calibration of this little sinusoidal RF generator...



Above — When connected to the Barebones PM, I calibrated the output power of the sine wave generator by tweaking the trimmer pot to give an output of 1.60v; the same Port B voltage yielded by the CMOS RF calibrator.

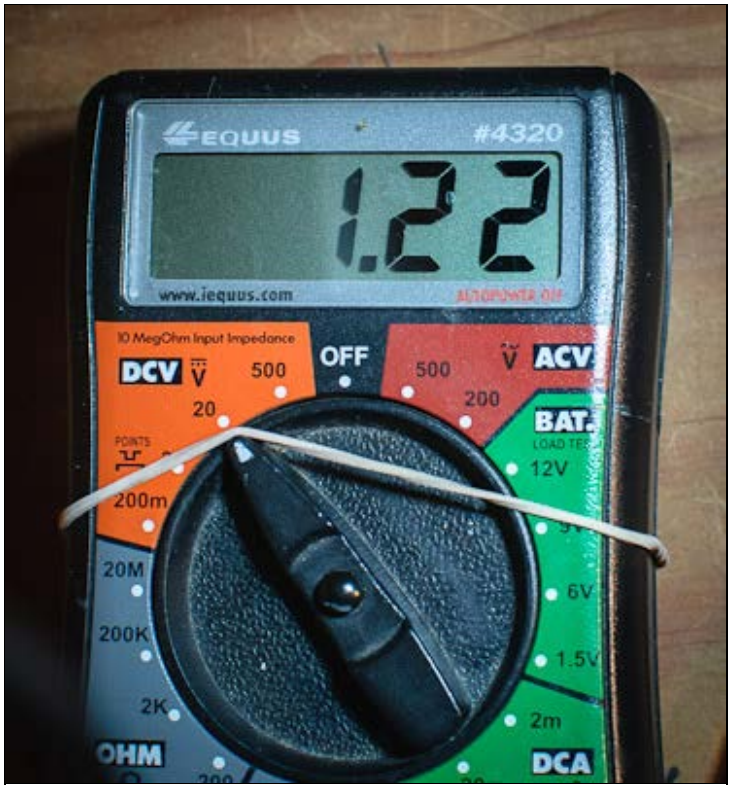


Above — The 50 Ω terminated 'scope verification of my calibrated 10 MHz sine wave signal source: 200 mV pk-pk = -10 dBm. Wow, thanks for this Bob!

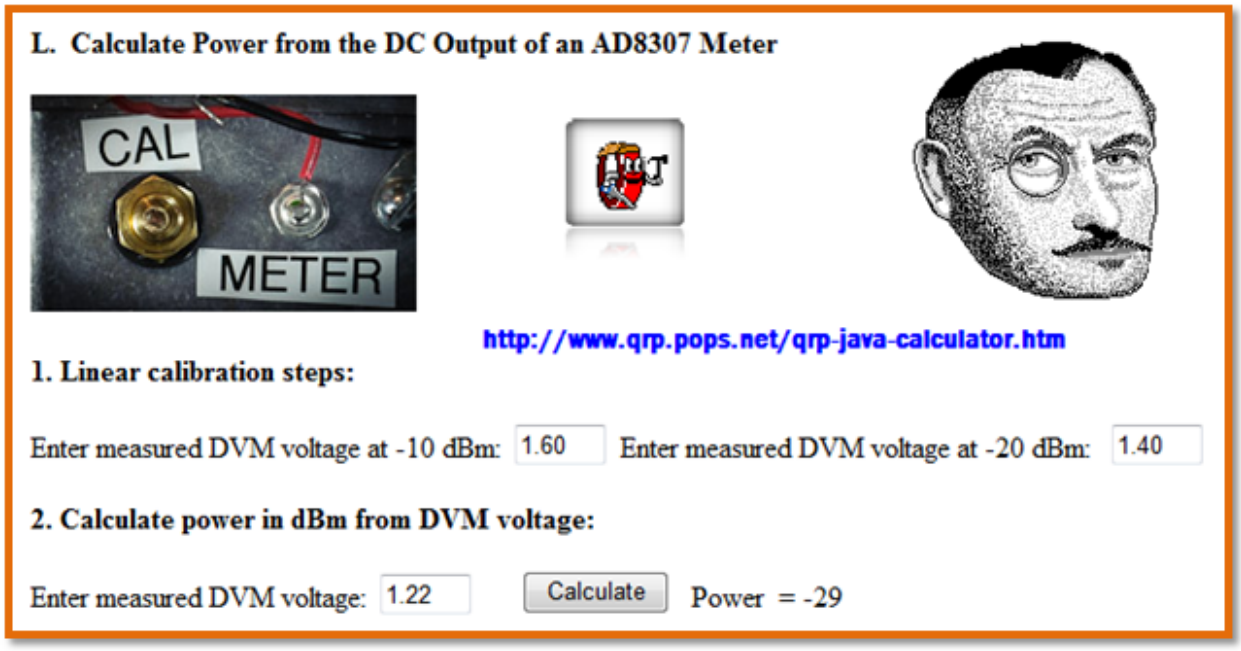
Above — The Port B voltage when a 10 dB attenuator pad was connected to the calibrated -10 dBm sine wave signal source: Measured power = -20 dBm. Notice the 20 mV/dB power drop — right on specification.

Refer to the section titled **<How to Measure Power at Port B>** Recall Wes' calculation needed 2 points to set the log linearity. I chose -10 dBm and -20 dBm instead of 0 dBm and -10 dBm so the CMOS RF calibrator could be used to calibrate any sine wave signal generators on hand.

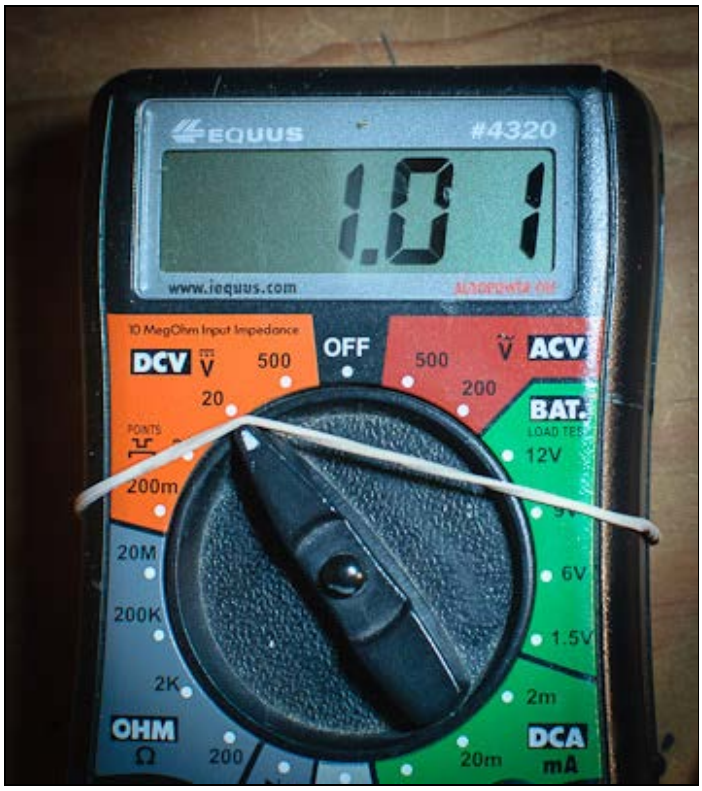
The -10 dBm calibration reference power serves as 1 of the calibration points in Applet L while we derive the other by adding a 10 dB attenuator pad to a *sine wave* signal source output port. ***Don't connect an attenuator pad to the CMOS square wave calibrator —*** error in the AD8307 arises. Why? Analog Devices mysteriously changed the AD8307 at some point after 2004 and altered its crest factor — this disallows us performing log linearity calibration with a square wave.



Above — A 20 dB pad connected to the -10 dBm sine wave signal source yielded 1.22v at Port B in the Barebones PM. See the power calculation with Applet L below:



Above — The calculation of the output power with a 20 dB attenuator pad on the sine wave signal source output. We're off by 1 dB since it should have calculated a power of -30 dBm. Could this be error caused by my attenuator circuit, or non-linearity by the AD8307 PM, or a bit of both? We have to live with such problems. Still, the Barebones power meter seems quite impressive for a simple circuit that I scratch built and calibrated in about 25 minutes.



Above — The Barebones PM DC output voltage with a 30 dB attenuator pad on the output of my -10 dBm, 10 MHz sine wave signal generator: Measured power = -39.5 dBm. Again; a very good — but not perfect power meter.

A Barebones power meter allows builders with modest equipment to measure power, gain/loss, and with a return loss bridge, return loss at HF. I wish I owned this little gem back when I started out.

Richard, "Dick", N4HAY, posted some great notes on his blog. I recommend following his blog since he really digs deep and likes math — thanks Dick!

Best!

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3. A Basic RF Workbench

Since January 2012, a handful of readers asked what I consider a good basic *RF* work bench. Again, I'm just an amateur hobbyist, so my opinion might show my ignorance.

A stand-alone 50 MHz oscilloscope with at least one 10X probe. *More bandwidth if you plan to work above 50 MHz*

If no oscillocope, an AD8307-based power meter *The modern version of the diode RF probe*.

3-4 50 Ω coaxial cables with BNC connectors; a 50 Ω scope feed-through terminator, *1-2* 50 Ω BNC port terminators, a though-connector and some BNC connectors to solder onto temporary circuit boards or mount in a chassis.

A homebrew return loss bridge.

3, 6, 10 and 20 dB BNC connector equipped attenuator pads, or a step attenuator.

Signal generator(s) that cover most of HF; +/- VHF signal generators described in the next section.

AADE L/C Meter IIB. Click

12 volt regulated power supply good for at least 1 amp.

Digital multimeter. *I use 2 and keep 1 set up for current measurement only.*

Frequency counter: homebrew or commercial. *I ran a 40 year old, ovenized, accurate HP counter until 2012.*

With these devices, as possible, you can work in a modular, 50 Ω environment and measure gain or loss in dB, return loss in dB and absolute power in dBm. Starting small and expanding your bench around 50 Ω input and output impedance devices will provide a lifetime of challenge and excitement in RF design.

Later, the big toys can follow: spectrum analyzers, VNAs, commercial signal generators and other lab quality stuff.

Equipping an RF bench presents quite a financial burden. I started small and slowly added pieces over time. Many pieces such as my L/C Meter IIB were gifts for holidays or my birthday. Other pieces were old, inexpensive equipment that I restored and calibrated.

Above — RF tools of the trade. We're RF experimenters! As scratch homebrew builders, gear like BNC, SMA and through-connectors, 50 Ω terminators and inline attenuators lie scattered on our benches; our fodder. Alternate photo.

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4. VHF Signal Generators

Having only started at VHF in November 2011, my knowledge suffers, however, a search for accessible, affordable, good quality VHF signal generators disheartened me. Ten year old or newer signal generators covering the VHF band work up to several GHz and cost a small fortune.

Lamenting old timers often recommend the vaccum tube HP-608 series that covered ~10-480 MHz. These heavy, glowing beasts sometimes come up in estate sales or on eBay for $200-400.

Then, too, the HP8640 series seems attractive, however, they are full of decaying parts. Ken Kuhn and others restore old HP gear as a hobby and this direction certainly gives us a valid option.

I've investigated 1 or 2 new, low-cost, commercial signal generators that work into VHF, but they failed to excite me; especially after I downloaded the schematics and sat in disbelief over their poor design.

Some minimum commercial signal requirements might include stable, linear tuning, a metal chassis, 50 Ω output with a return loss greater than 20 dB and low harmonic distortion at all frequencies. Like the rest of our lives, our budget usually determines what we buy.

I decided to build my own VHF signal generators and document them on this web site. I've learned that home building signal generators between 50 and 200 MHz requires skill and care, but can be done.

What about digital clocks?

At VHF, DDS spurs get extreme as you get closer to the maximum clock frequency .

The Si570 looks intriguing, however, still requires an MCU + components, I haven't read any lab quality evaluations of the Si570 as part of a engineer-grade VHF-UHF signal generator and if you know better, please email me.

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5. L - C Meters

If you search for opinions about which L/C meter to get, you'll find an abundance of super write-ups including those that cover measuring with Kelvin probes, SMD tweezers; or statements suggesting that if you really need maximum accuracy, purchase a VNA. I encourage you to research this yourself and find the best L/C meter for *your* bench. Here are my 2 cents worth of opinion and please remember — I'm often wrong.

I use an AADE L/C IIB meter to test inductors and capacitors for HF and even some VHF work. Yes, the AADE L/C IIB doesn't measure large-value electrolytic caps and so forth, however, considering the cost versus performance — it's accurate enough for the popcorn RF workbench.

The AADE L/C meter uses the method described by William Carver, W7AAZ in an article called The LC Tester published in Communications Quarterly, Winter 2003 . EMRFD page 7.12 briefly examines Bill's circuit and shows his original oscillator along with an extended range Colpitts oscillator designed by Wes. For brevity sake, I'll just discuss inductance measurement with Carver-style meters. With care, an inductance resolution of 5 - 20 nH might be realized with such a device.

We normally don't consider that our inductor is actually a network with L, a parallel C and losses that might be modelled as R in series with the L, or a R in parallel with the C depending on our model. The inductor also exhibits a self-resonant frequency and for our design purposes, we usually *ignore* all these details and just consider it a "pure L". My L/C meter's oscillator runs from a few tens to a few hundred KHz and generally lies below the self-resonant frequency of the inductors I measure with it. I've learned by sweeping/analyzing my completed filters, that as long as you avoid the coil's SRF, *the low frequency* Carver-style meter proves a stalwart inductance meter for most HF and some VHF applications.

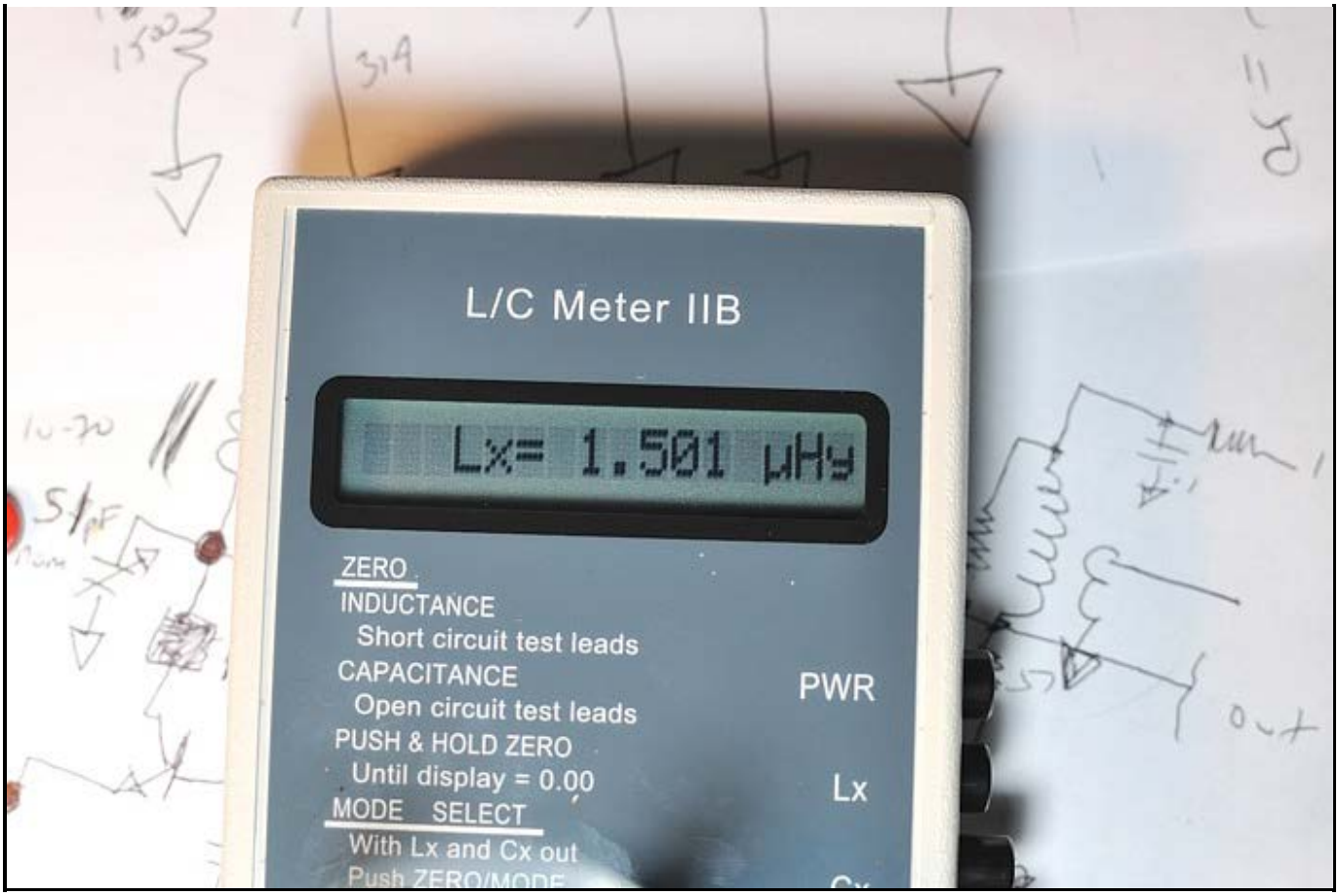
Often, we popcorn builders want to make a filter, an oscillator, or a pi, or L match and we apply software or tables to calculate the L and C values needed to resonate our filter tank(s) or matching networks, or to synthesize a low-pass filter. These math-driven programs/tables assume the *pure* "pixie dust" L described eariler — disregarding the stray C and R. So getting all worked up about whether our coil is 4.50 or 4.59 microHenries seems moot.

Further, we man-handle our inductors [changing the inductance somewhat] into a breadboard laden with 5% (or greater tolerance) capacitors, copper board pads/paths that exhibit C, active devices and so forth. Then, too, we connect these resonators, or filters to other blocks with sometimes reactive ports, plus or minus shielding. Despite all these variables, miraculously, we make the filter with our "measured" L work!

As we move up in frequency, the effects of stray L effect magnify and at some point our filter networks may behave poorly.

For band-pass filters with Carver-style device measured inductors, we need only adjust each trimmer capacitor to get the highest possible peak-peak voltage, or RF power with our filter between a signal generator and a 50 Ω terminated scope, power meter *respectively*, or whatever. After sweeping these peaked filters, rarely do I need to compress/expand, or add or remove windings to tweak the L get the desired filter response when the filter input and output ports are well matched.

In the case of single frequency matching networks like the L-match, we might need to tweak up or down the L to derive strong port matching. In all cases, our software and the L/C meter can get us into the ballpark, but in-situ bench measurement with other instruments will garner the home run.



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6. Bob, K3NHI — RF Power Meter Follow-on

Bob built a follow-on detector and power meter to his 2002 QEX power meter *[Reference 3 in Section 7].* Bob gave me the green light to share his creation on my site. Thanks awfully Bob!

Click for the schematic in pdf format. Per typical K3NHI fashion, it's laden with trimmer pots allowing precise calibration — Bob's stuff contains loads of tweaks and wiggles! The new PM offers more flexibility + features including bigger battery supplies that won't quickly die when you fail to turn it off. He also included a means to measure battery voltage on the analog meter. Click for a photo showing the batteries. Low noise, high-speed, rail to rail, CMOS op-amps for IC1 and IC3, allow you to capture PEP during sideband transmitter measurement with appropriate attenuation.

The TS922 op-amp might be hard to find in DIP since that package went obsolete, however, it's available in SMD. If you can't get any TS922, likely other modern, high speed, rail to rail, CMOS op-amps will work fine — consult datasheets to ensure you meet or beat the TS922's performance for IC1 and IC3. Some of the latest design op-amps offer truly sublime specifications and evoke joy in our breadboards.

Bob critically *isolated* the entire RF sub assembly **outlined in green on the schematic** with metal shielding and feed through capacitors for the B+ and DC output. This helps ensure stable and accurate AC voltage measurement. Click or click or click for more of Bob's photographs.

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7. References

[1] Roger Hayward, KA7EXM — A PIC-based HF/VHF Power Meter, QEX May/June 2005

[2] Wes Hayward, W7ZOI, Bob Larkin, W7PUA, Simple RF-Power Measurement, QST, June 2001

[3] Bob Kopski, K3NHI — An Advanced VHF Wattmeter, QEX, May/June 2002.

[4] Bob Kopski, K3NHI — A Simple Enhancement for the Advanced VHF Wattmeter, QEX, Sept/Oct 2003.

[5] Bob Kopski, K3NHI — A Simple RF Power Calibrator, QEX, Jan/Feb 2004 **+** Tech Notes article, QEX, May/June 2010.

RF — Test and Measurement

VHF FM

Yes - I could purchase a FM radio-in-an IC for $2.00 and be done in 35 minutes, but what would I learn?

Repository for FM superhet receiver experiments conducted from 2012 to 2014.

1. 10.7 MHz IF Filter Experiments
2. A Basic Colpitts VCO
3. DC-DC Converter for VCOs
4. Supplemental Page #1 - it's time to make some receivers
5. Miscellaneous Photos or Figures

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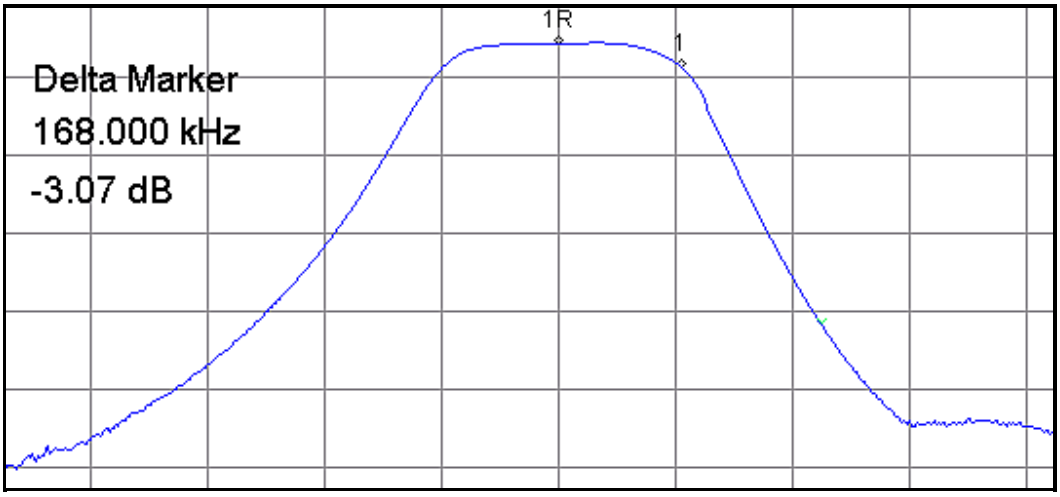
1. 10.7 MHz IF Filter Experiments

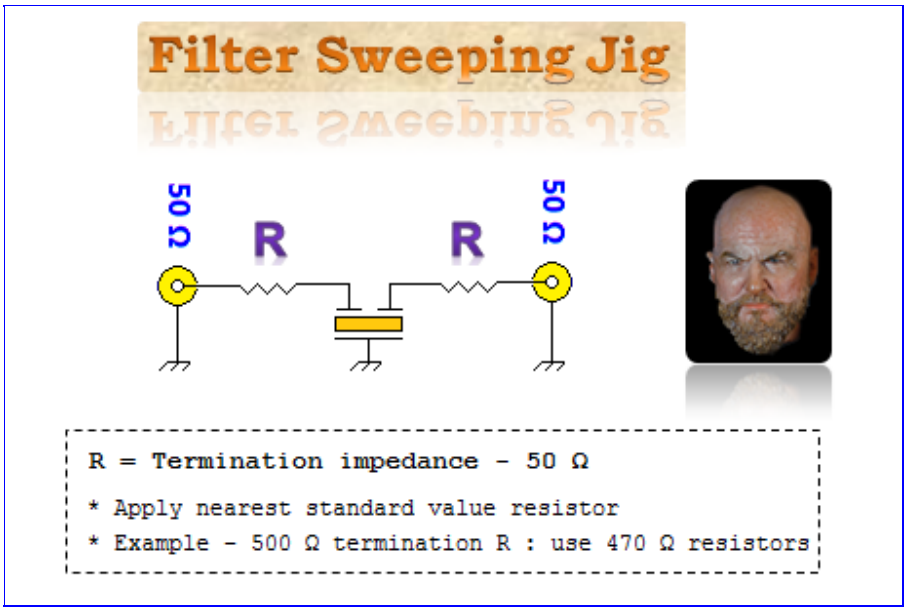


As a FM receiver design newbie, I read about and experimented with some 10.7 MHz IF filters to learn common practices, what's available and which measurements might help me to reach my goals. Over time, I've collected a variety of crystal and ceramic filters for hopeful future work. Click for 2 exotic examples. IF filters might be purchased at Ham festivals, surplus electronic parts stored and/or online. Prior to paying for a

filter, I've found it useful to politely request a sweep of the filter, or, better yet, perform this task myself. To sweep a filter in your lab, you'll need a tool such as a spectrum analyzer with tracking generator, a VNA, or some other analog/digital sweep system.

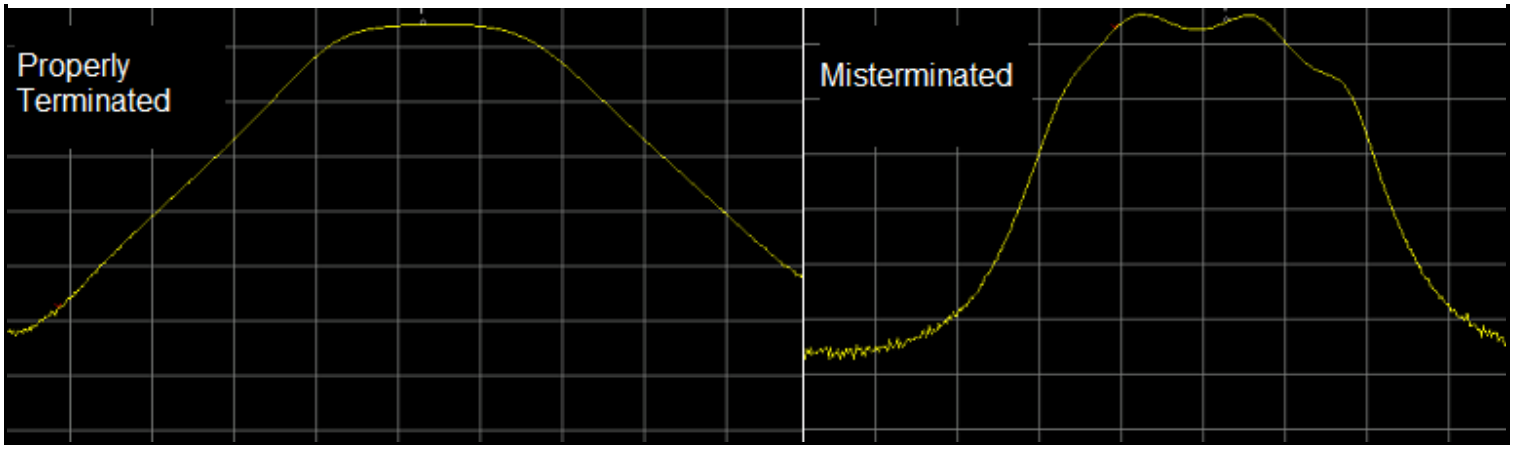
I'll homebrew some crystal filters for narrow band FM in future installments — I ordered some 20 MHz xtals.

Sometimes a filter in your junkbox will state the IF and perhaps the 3 or 6 dB bandwidth, but not the input/output port termination impedance. How do we determine this impedance? I've learned we can figure this out by testing differerent termination resistors with this simple test jig:

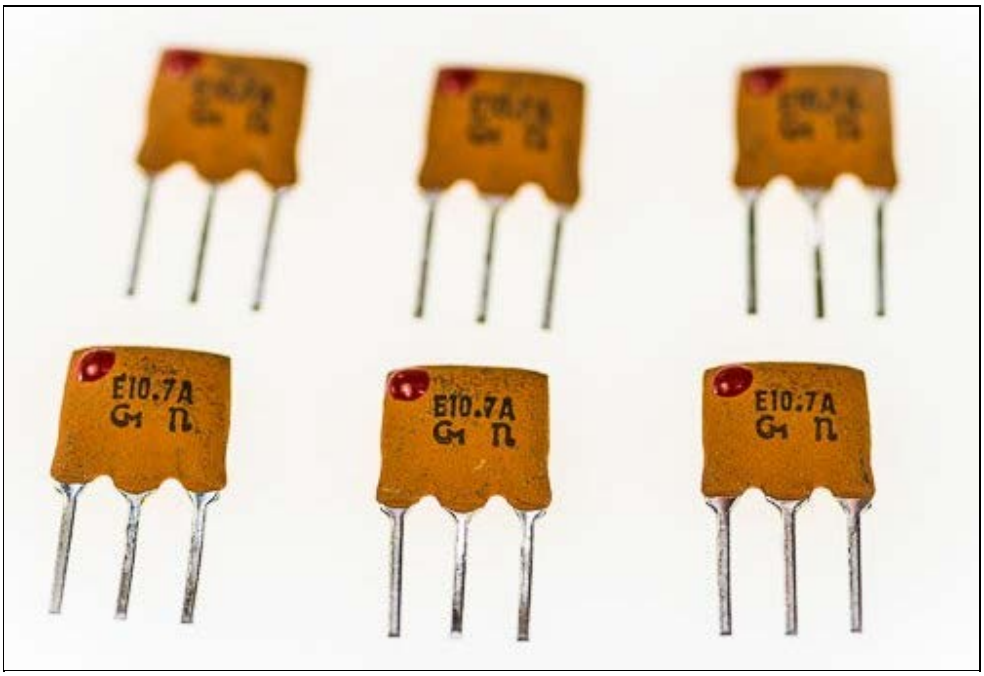


Above — A simple crystal or ceramic IF filter sweeping jig. Since the series resistors attenuate the signal, losses occur; but the shape should look clean with minimal ripple. Normally, we builders will also place (or switch in) 50 Ω attenuator pads on both the signal source and detector within our sweep system to buffer impedance mismatch. Comparisons of this simple jig with more precise and complicated matching methods suggest that for many filter sweeps, it might work fine.

Above — My test jig with a Murata ceramic filter soldered in-situ. Keep the resistors *close* to the board. I've pretty much moved to SMA connectors in my lab: they're cheaper than BNC, plus we can buy a wide variety of quality 50 Ω patch cables donning various connectors for low cost. For example, a 30 cm cable with a male BNC and SMA connector on either end.



Above — A poor termination may result in improper bandwidth and ripple — easy to spot in this trace. Click for a trace from a 'gone bad' instrumentation crystal filter: 10.7 MHz @ 30 KHz with 2200 Ω Z in/out. Not really usable with ~ 10 dB ripple.

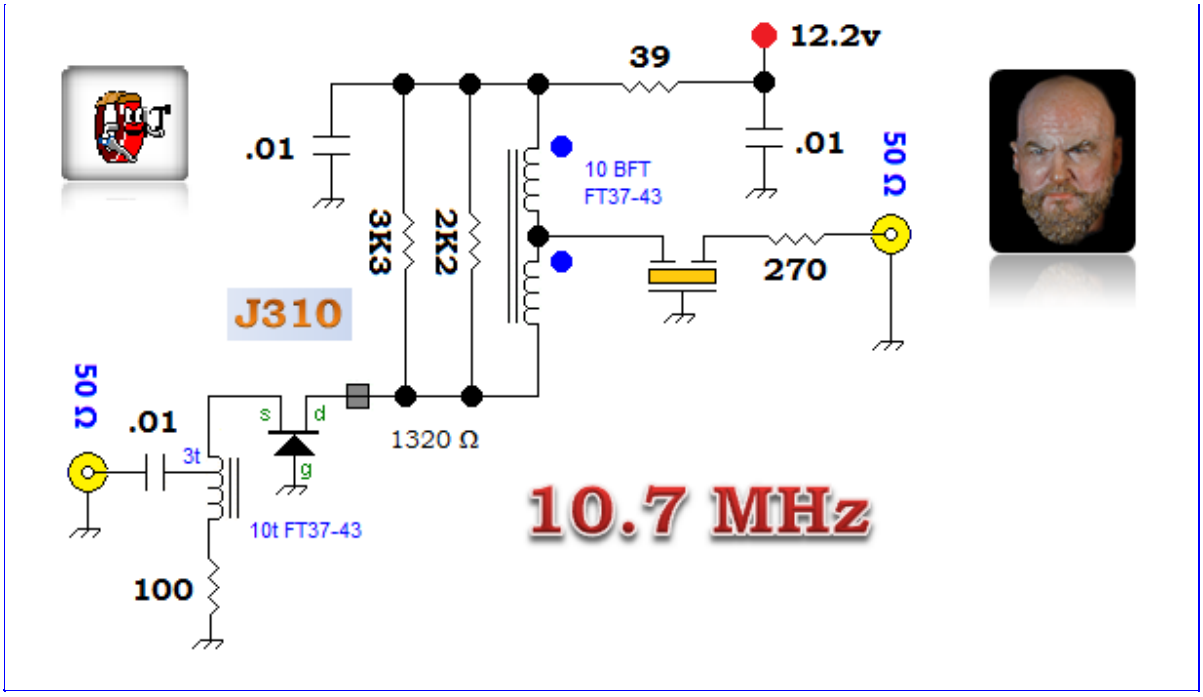


Above — Older 280 KHz Murata 10.7 MHz IF filters purchased long ago. Low cost = their main attraction, although they too will suffer total obsolescence and a price increase.

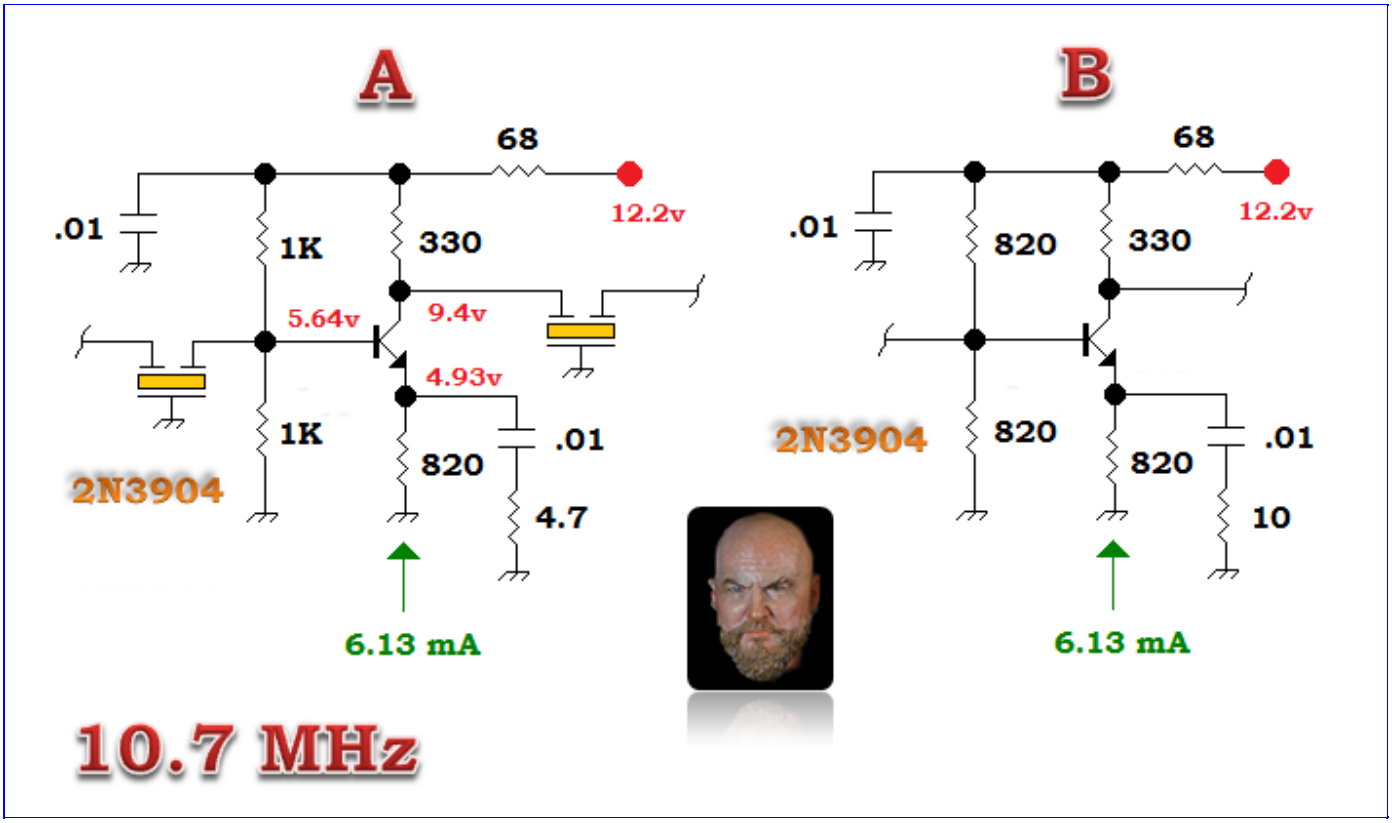
I bought some newer, lower insertion loss ceramic filters in the following bandwidths: 280 KHz, 230 KHz, 150 KHz, 25 KHz and 20 KHz. Check their datasheet — most Murata ceramic filters require a 330 Ω termination (preferably resistive) and I keep a filter sweeper jig with 270 Ω resistors as a regular bench tool. 280 KHz was a popular WBFM filter bandwidth in many older high-end FM receivers including my 1980's T-85 Yamaha receiver; my benchmark FM receiver.

Many of us hopeful FM builders, smitten by modern digital gear, fail to recognize the fantastic design achievements made by FM receiver engineers back in the day. All those air-variable, ganged band-pass preamp stages, low noise amplifers and often incredibly complicated and great sounding FM multiplex circuits just blow me away. Perhaps I'm a hopeless analog nostalgic? My T-85 sports 5 ceramic filters [280 KHz and 230 KHz B/W Muratas] and the narrow filters are listener switchable for narrow band Dx.

Ways to Match These Filters with Amplifers

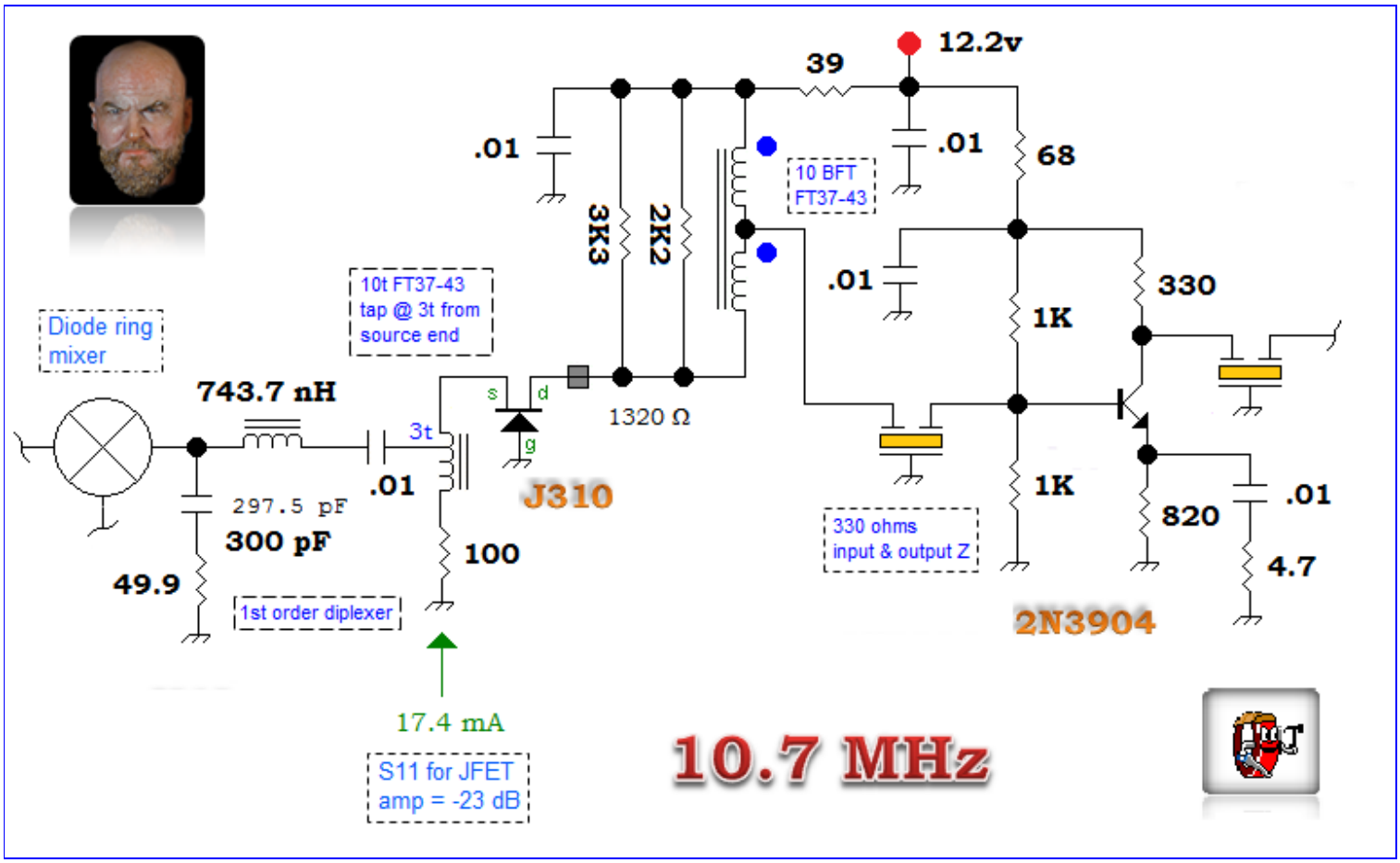


Above — A common gate JFET amplifer drives a 330 Ω ceramic filter. I placed 2 resistors in parallel to get the needed shunt R of 1320; my 2 resistors measured 1316 Ω. The bifilar transmission line transformer provides the 330 Ω Z to drive the filter. Details of the JFET amp come in a later schematic, but the input return loss at 10.7 MHz = 23 dB. I swept this circuit and it looked similar to the tracing with the same ceramic filter in my 270 Ω filter sweeping jig. This particular filter exhibited 6.4 dB of insertion loss.

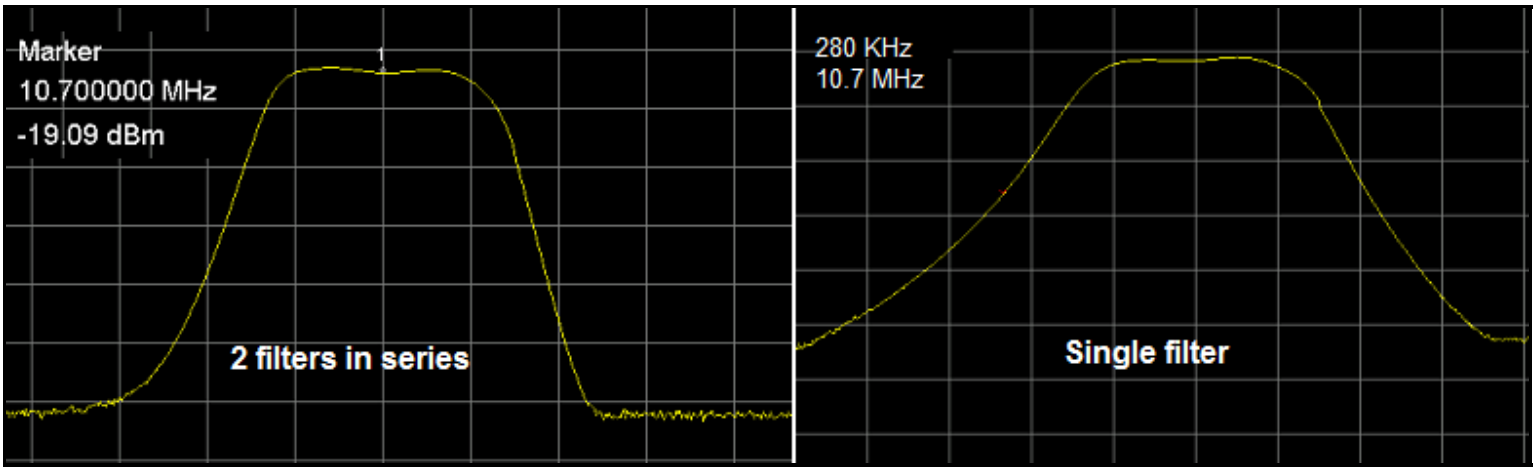


Above — 2 versions of a BJT amp with a 330 Ω input and output impedance. If you read schematics of good FM receivers, often the designers drive the filter with a 330 Ω collector resistor. Click for an example. By keeping the bias and degeneration resistors low and the current moderate, an amplifer with 330 Ω input Z is easy to design [although the input Z will vary with Beta].

I felt surprised that version **A** exhibited a voltage gain of 11.7 despite those low bias and collector resistors. You can stick a filter on either side as shown. Murata recommends a buffer amp between cascaded ceramic filters and you'll see this often in FM receiver schematics from the 1980s or so. Resistors provide wideband termination. Version B is the same amp with a little more degeneration to lower the gain and serves as a design example. I've got the procedure documented here under '*Calculating the input resistance of a common emitter stage'.*

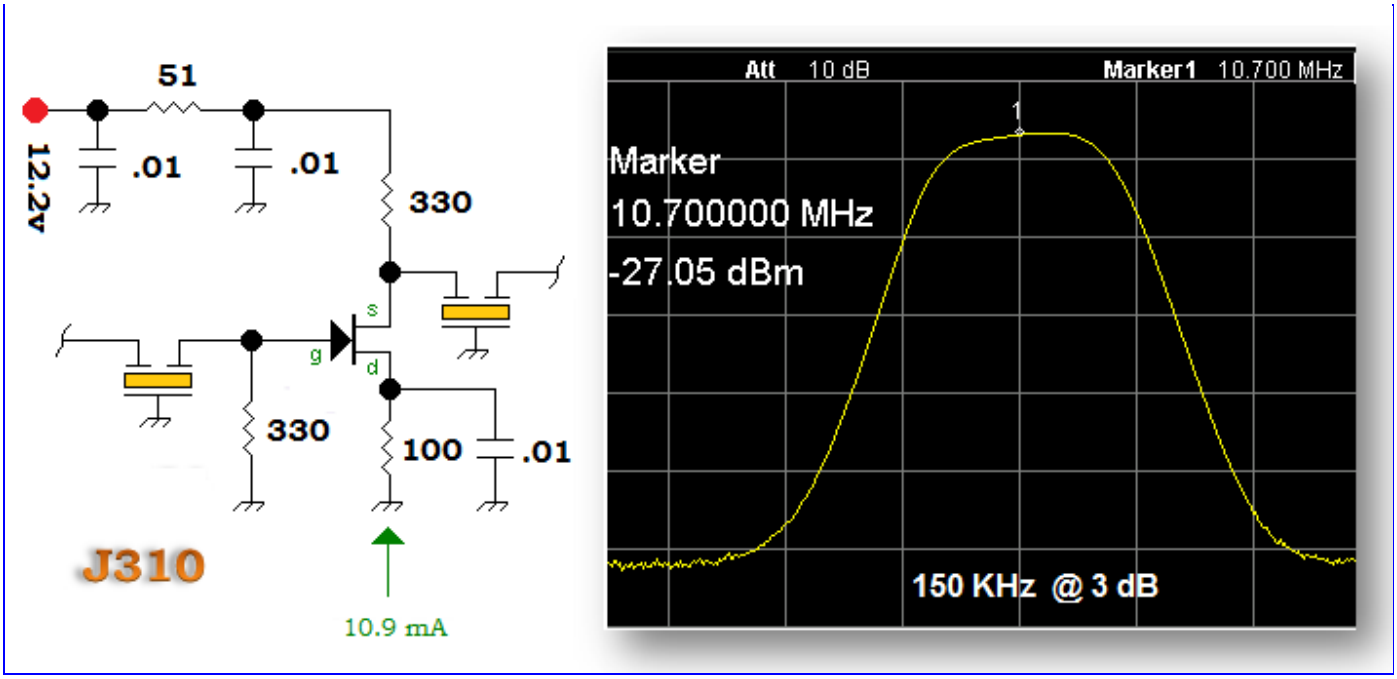


Above — An IF block using the designs shown earlier. I terminated this stage with a 270 Ω resistor and of course removed the mixer and diplexer. 2 sweeps lie below. I'm tempted to tune the JFET drain and couple the transformer with a few links as needed to get a 4:1 impedance ratio. Anyhow — food for thought.



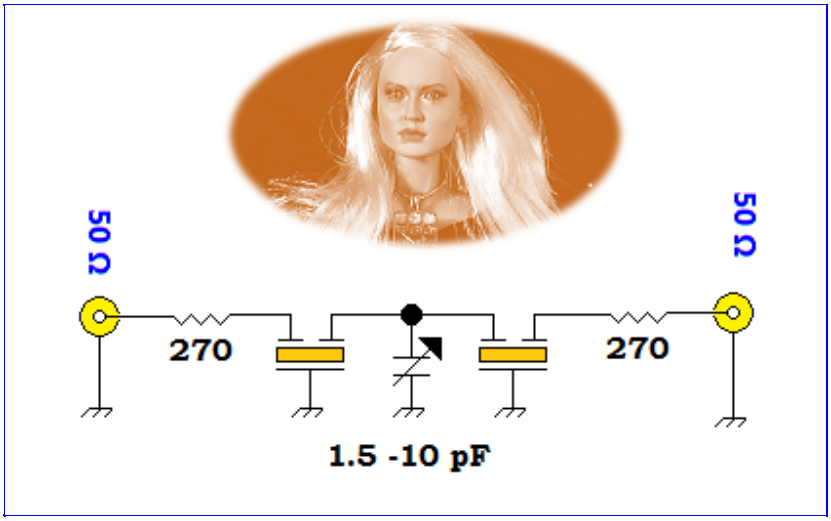
Above — A sweep of the IF block shown above left sans the mixer + diplexer (*two* 280 KHz ceramic filters). On the right lies the trace of the common gate amp driving a *single* 280 KHz filter with its output terminated with a 270 resistor. The advantages of 2 cascaded filters seems apparent, although the slight downward dimple at the center frequency might represent some capacitive loading at the output of the common gate amplifier.

I built a number of other amplifers and swept all of them Click or click for 2 early examples that use active devices instead of series matching resistors on the output. In these circuits, R Term was changed and then the circuit was sweeped. The tracings looked good.



Above — I built Brian, K6STI's nJFET IF amp. He used it to offset the losses associated with 2 ceramic filters. Click for Brian's fabulous website. I placed 150 KHz 3dB BW filters before and after a J310 and swept — my circuit exhibited a 2 dB net loss which seems quite reasonable.

The 330 input resistor is a load/termination on the input filter and will dissipate some energy and lower the AC input voltage to the gate compared to the usual high Z input resistors we apply in our JFET common source amps — from open circuit to full termination would incur a 6 dB voltage drop. Still, for simplicity versus performance, Brian's circuit looks hard to beat.



Above — 2 ceramic filters in series. I added a small trimmer between the pair in hopes to mitigate any filter skirt distress or ripple. Click for a tracing with and without the trimmer capacitor. You might experiment with the filter coupling and the filter block termination impedances to better their skirts and passband The losses of the above filter block may reach 12-14 dB.

If you don't have a sweep system, I was able to crudely test the amplifers + filters with my 10.7 MHz signal generator and a DSO.

Resistance Bridge

If you go with a BJT IF amplifer, it's possible to measure the input impedance with a bridge and tweak the emitter current and/or degeneration resistor to get very close to a 330 Ω Zin. I keep a drawer with through-hole resistors rated between 1 and 10 Ω for tweaking my emitter resistor values to change series feedback in my common emitter amps.

I first designed a simple 330 Ω bridge for measurement with my DVM. It worked, but the null lacked the depth and resolution we need. Later I improved the sensitivity by adding another coil and changing to a 'scope or SA detector, but after building EMRFD Figure 7.36, I abandoned my bridge. Figure 7.36 just blew me away. The null of a 330 resistor was only a few 10s of microvolts during calibration.

I placed a small 500 Ω pot in parallel with a 120 Ω resistor for the variable resistance. After some basic testing, I calibrated it with a 330 Ω

resistor; adjusting the pot for the deepest null and just left it there for testing my 330 Ω IF amps @ 10.7 MHz.

I plan to make Figure 7.36 for VHF and maybe UHF with chip caps plus a small screwdriver adjustable trimmer pot [to get the lowest possible L] calibrate it and make it a part of my test bench arsenal. After getting a null, we measure the pot's resistance with an ohm meter to learn the impedance at the ? port

Considering that our predecessors measured just about everything RF with a bridge, this little circuit suddenly become relevant. A series L and C "add-on" circuit shown as Figure 7.39 may be placed in series with the ? port and device under test to deepen the null in the face of reactance. Bridge circuits form the very essence of RF measurement. Yes Bobby, we can measure impedance without a VNA.

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2. A Basic Colpitts VCO

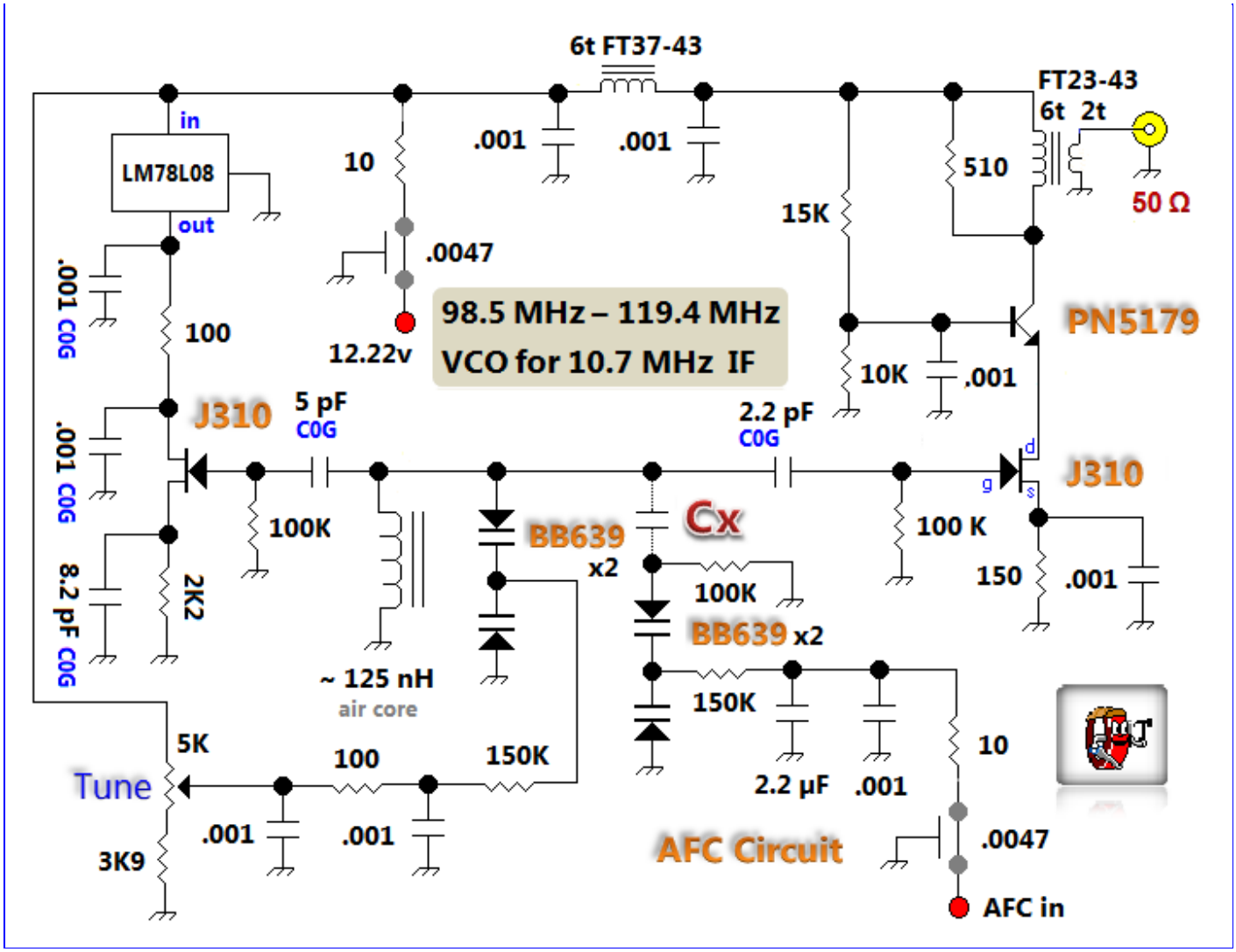
Above — My completed Colpitts VCO. *I installed the unlabelled, left-sided pot in case a potentiometer is required for future AFC circuitry changes. It's not hooked up.*

I reviewed some 1970's FM receiver schematics to learn that before PLL-locked VCOs dominated, often Colpitts VCOs were locked onto a strong frequency with Automatic Frequency Control (AFC). Local oscillators tanks often employed a inductor plus an air variable capacitor that tuned from ~77 to 119 MHz with a varactor for AFC. All the tuning and front-end filter air variable capacitors were ganged together and I'm sure alignment took some skill.

Some VCOs tuned with varactor(s) instead of an air variable cap — this is what I wish to do. Varactor tuned VCOs usually suffer more thermal drift than air variable capacitor versions.

AFC compensates for VCO thermal drift by a seperate varactor with its control voltage line DC coupled to the FM detector through an R-C low-pass filter. Any difference between the VCO frequency and the desired FM frequency produces a proportional DC voltage. The DC control voltage changes the oscillator to the desired frequency by re-tuning the AFC varactor within this feedback loop, albeit over a limited range. AFC is unsuitable for weak signal DXing, since it may pull the receiver onto a strong adjacent signal. Many 1970's FM receivers supplied an AFC defeat switch.

I remember 1 old FM receiver in my parent's home that stayed locked on 1 frequency for years thanks to AFC.



Above — The schematic of my version of a JFET Colpitts VCO (with AFC) that lacks the standard gate to source feedback capacitor; the intrinsic capacitance from the J310 gate to source provides the feedback needed for oscillation.The 8.2 pF bypass cap was determined on the bench — too little, or too much C decreases output voltage, or snuffs out the oscillator.

I just couldn't bring myself to make a VCO with a BJT, since on my bench at least, they suffer more thermal drift than JFET-based oscillators. I built with a mixture of SMT and hole-through capacitors and resistors. The anti-parallel arranged hyperabrupt varactors were found on eBay. Click for a rear photo of the project chassis. The gold colored jack is an SMA connector.

I bench designed this VCO and it took many hours to find the correct amount of L and C for the resonator to give a low distortion, sine wave output across the ~21 MHz tuning range. This meant soldering in and removing these tank components frequently. Click for the lowest frequency output. Click for the highest.

In the example local oscillators I reviewed, the engineers made no attempt to level off the signal that normally increases in AC voltage as you increase frequency. I also ignored levelling. Presumably the designers didn't worry with leveling the oscillator output in their superhet receiver as long as the output voltage sufficiently drove the mixer into complete switching. Levelling would add cost and complexity. This isn't a lab grade RF signal generator — that's for sure.

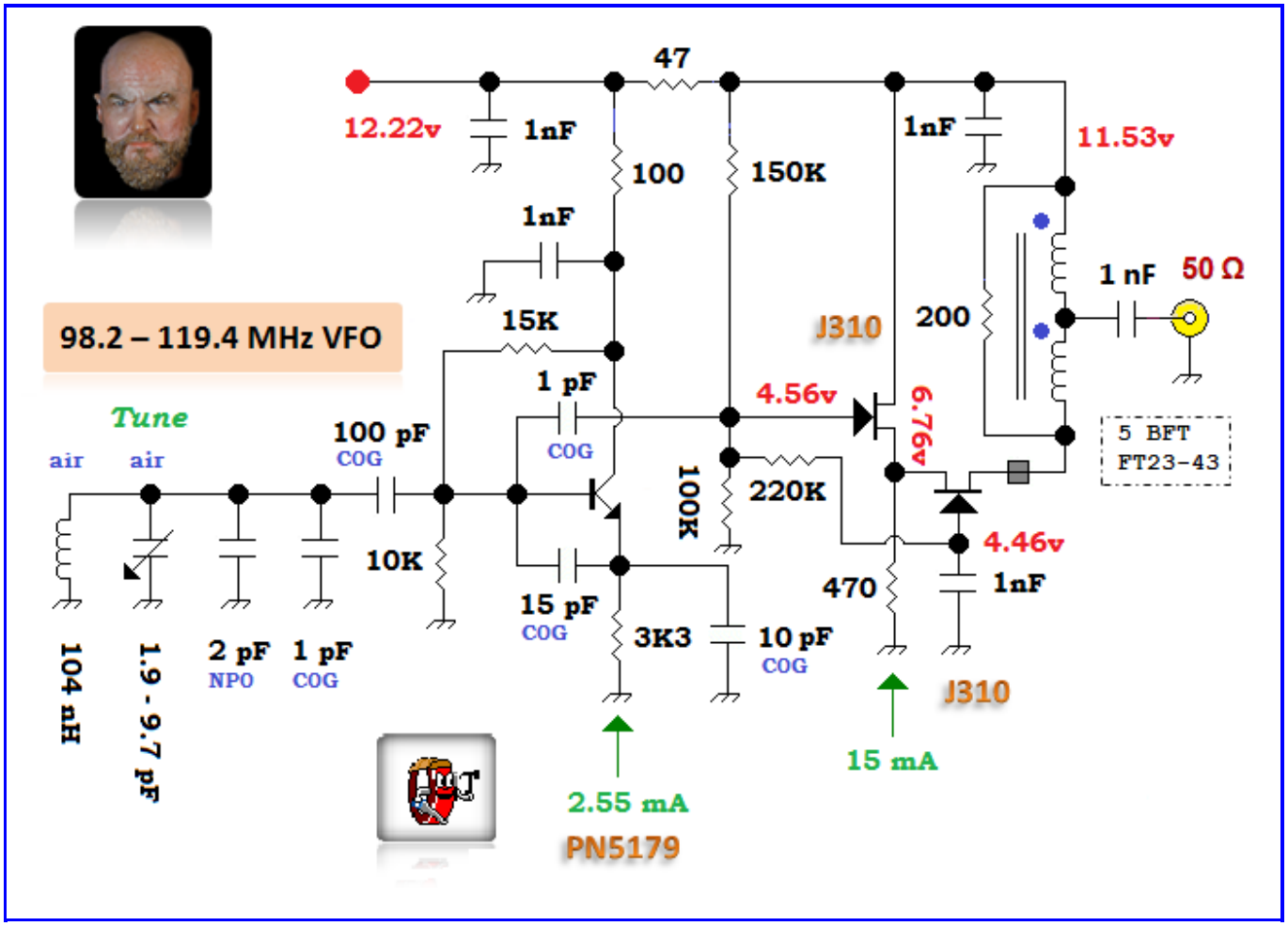
At present, the AFC varactor pair is disconnected since I won't know how strongly to couple it with Cx until I have a working detector. Also I will need to experiment to determine the best R-C time constant for the low-pass filter; likely the 2.2 uF capacitor will need an increase in value.

With the 3K9 Ω resistor under the 5K tuning pot, I keep at least 5 VDC on the tuning varactors or the VCO would stop running as I tuned the pot towards CCW. The coil = about 3 turns of 16 gauge wire on a 5/8 inch bolt. (Despite Canada going metric in ~1975, they still sell nuts and bolts in inches at our hardware stores). The stiff wire prevents the inductor from turning into a "microphonic" spring when the VCO is bumped. Click for a photo. The nominal L = ~ 125 nH, although I bent and manipulated the coil so it sat attached to the copper clad board with no tension and then squished or expanded the turns to establish my lower band edge.

In many FM receivers, either a single or balanced dual-gate MOSFET mixer was driven by a high impedance buffer/amplifer. If I mix with a 2-gate MOSFET, I'll insert a common gate JFET amplifier on the IF strip to boost the LO output impedance and AC voltage.

The feedthrough capacitors are 0.0047 µF - they were on sale so I bought them. To prevent a parasitic high impedance when placed in parallel with my standard 0.001 µF bypass caps, I placed a series 10 Ω resistor.

I enjoyed this crazy design; trying to replicate a relic, but popular local oscillator idea from decades ago. Let's hope I did it justice. Perhaps future VHF stuff on the FM and even 2 meter band will involve an Si570 and PIC, Arduino or other microcontroller? This simple VCO will do for now. My greatest passion lies in designing and building the front end.



Above — A well-buffered, bench-module, high-side VFO I sometimes use for broadcast FM band mixing into a 10.7 MHz IF. The output at 98.5 MHz = -5.35 dBm, perfect for switching Gilbert cell mixers with a little padding or amplitude tweaking. Click for the output of the Colpitts only with a 10X probe @ 120 MHz. With care, you can see the second harmonic in the 'scope tracing — click for the SA tracing that shows the 2nd harmonic 27.5 dB down from the carrier. Click for a 'scope tracing with my MMIC bench module amplifier from VHF Veronica connected; the amp exerts some low-pass filtering that cleans up the signal somewhat.

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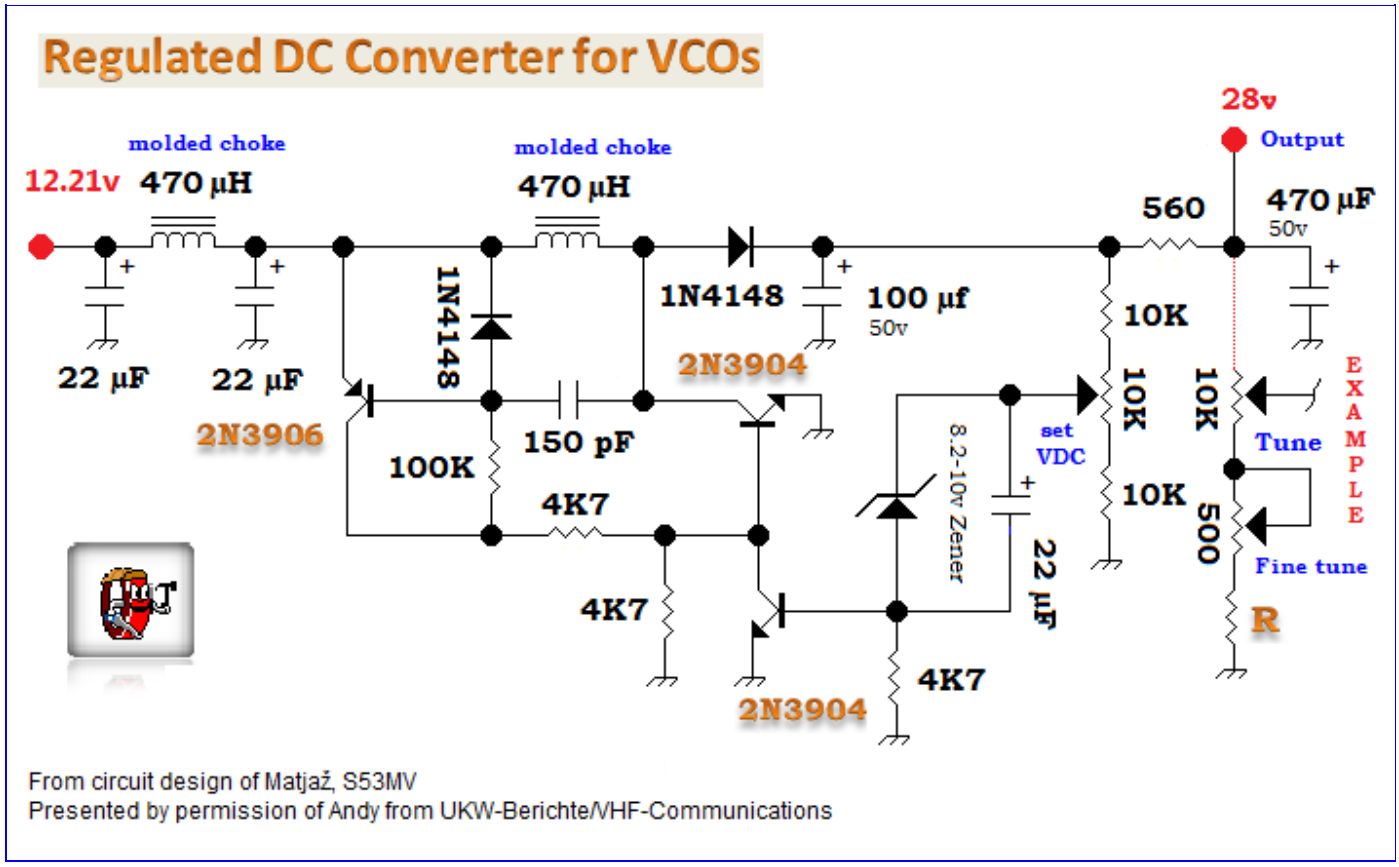
3. DC-DC Converter for VCOs

Until now, I ran a maximum reverse DC voltage of ~12 volts in my varactors. For wider VCO or L-C filter tuning, builders may chose 28 volt varactors such as the BB535 or BB149A and boost the 12v supply up to 28v with a DC-DC converter. Some build inductorless converters pulsed from 555 timers, or use CMOS voltage converters like the CL7662, or Si7661 to make a doubler. As an RF constructor, I like working with coils and built the following circuit:

Above — Bench module: 28v DC to DC converter. While containing no tuning control pot, my build places the zener diode regulator control potentiometer on the front panel to allow fast-tweaking of the output voltage from ~21-30 VDC depending on the load. Click for the breadboard photo.

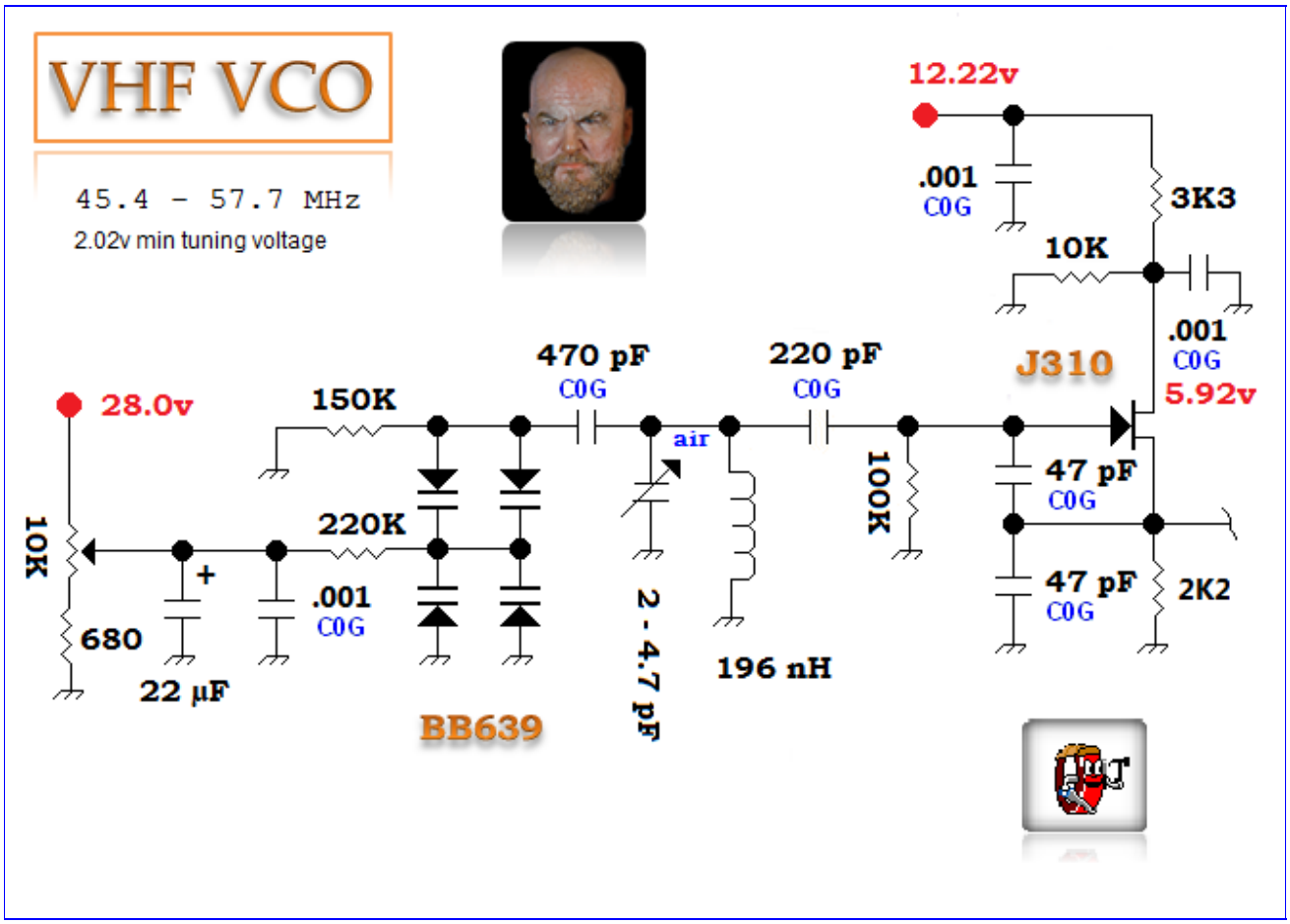


Above — My regulated 28v converter for varactor tuning adapted from a design by Matjaž, S53MV. I pulled this circuit from his *amazing* 2-part article with circuits that span from 11 GHz RF to DC. *See the reference articles below*. I filtered heavily and at switch-on, my circuit draws ~ 50 mA, but then drops to ~ 11 mA after the capacitors charge. The 10K *[set VDC]* trimmer pot allows you to dial in your desired output voltage and thus this converter may work over a wide range of DC power supply voltages.

The tuning control(s) might be a single potentiometer, or even seperate pots for tuning 2 different VCOs. In the above schematic, I show 1 possible tuning scheme: a 10K coarse tuning in series with a 500 Ω fine tuning potentiometer. R keeps some minimal reverse DC on the varactor(s) and is optional. Again, my bench module DC converter omits any tuning controls — these are built into the circuit containing the varactor diode(s).

The oscillator frequency varies slightly with the set output voltage. Click for a screen capture at ~32 VAC with a 10X probe placed on the PNP emitter. In another test, with no load, I watched the coil's magnetic field collapse and ring in this cool 'scope capture. This is why I love deep memory DSOs so much.

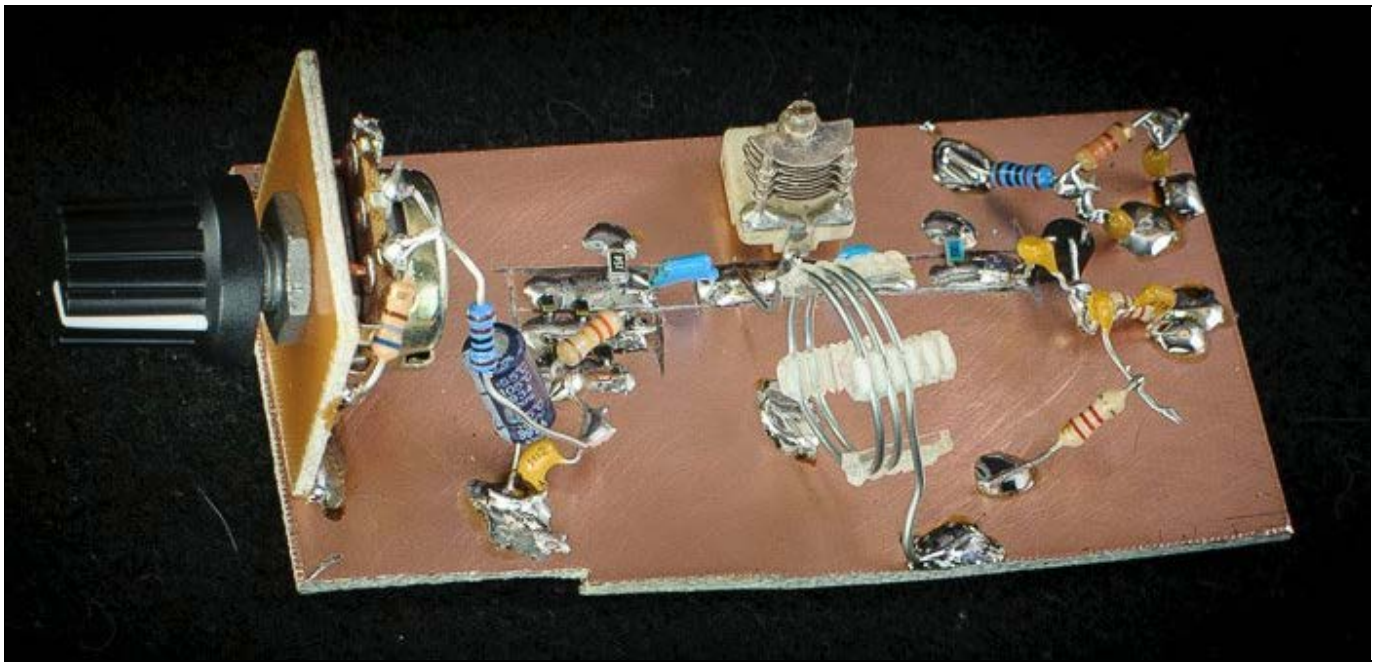
Above — Andy, G8ATD who owns VHF Communications magazine granted me permission to show the DC converter circuit. His magazine archives provide a treasure trove of useful circuits from VHF to Microwave and it's clear Andy passionately spent lots of time publishing the magazine until 2013, plus scanning and organizing the archived material.
Although VHF, UHF and microwave focused, much of the concepts and learning can also enrich your HF exploits.



Above — A "quicky" VCO thrown together to test the DC-converter. Click and click for the output with the tuning pot set to fully CCW and then

CW. The 680 R keeps about 2 volts on the varactors with the tuning pot set to CCW.

In the reference articles cited above, you'll find 2 HF-VHF Hartley VFO designs that tune over a 20 MHz span thanks to 28 volt varactors and careful design. In yet another UHF circuit, a 1 octave tuning span is realized with the author's specially designed VCO. Truly hardcore design from a great teacher — I crave exposure to the work of such authors.



Above — A photo of my "quicky" VHF VCO. 73!

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4. Supplemental Web Page #1

Click for the first supplemental web page.

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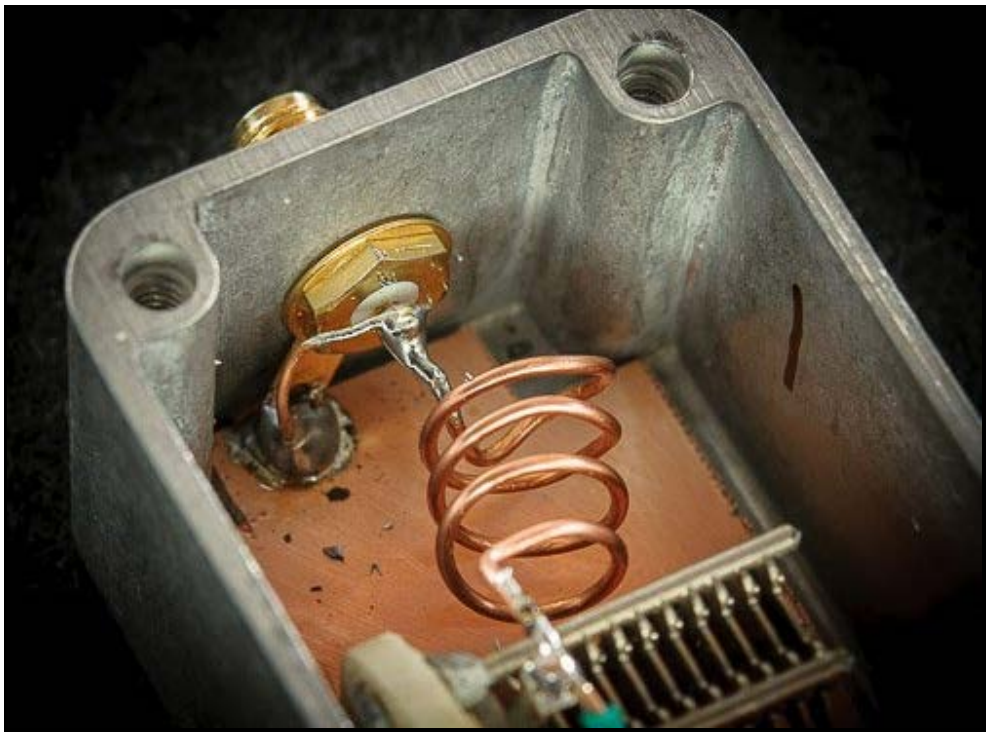
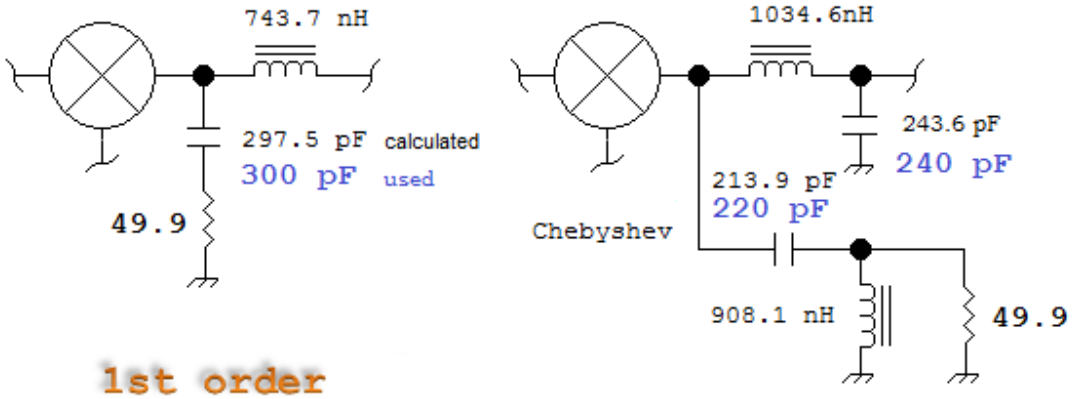
5. Miscellaneous Photos or Figures

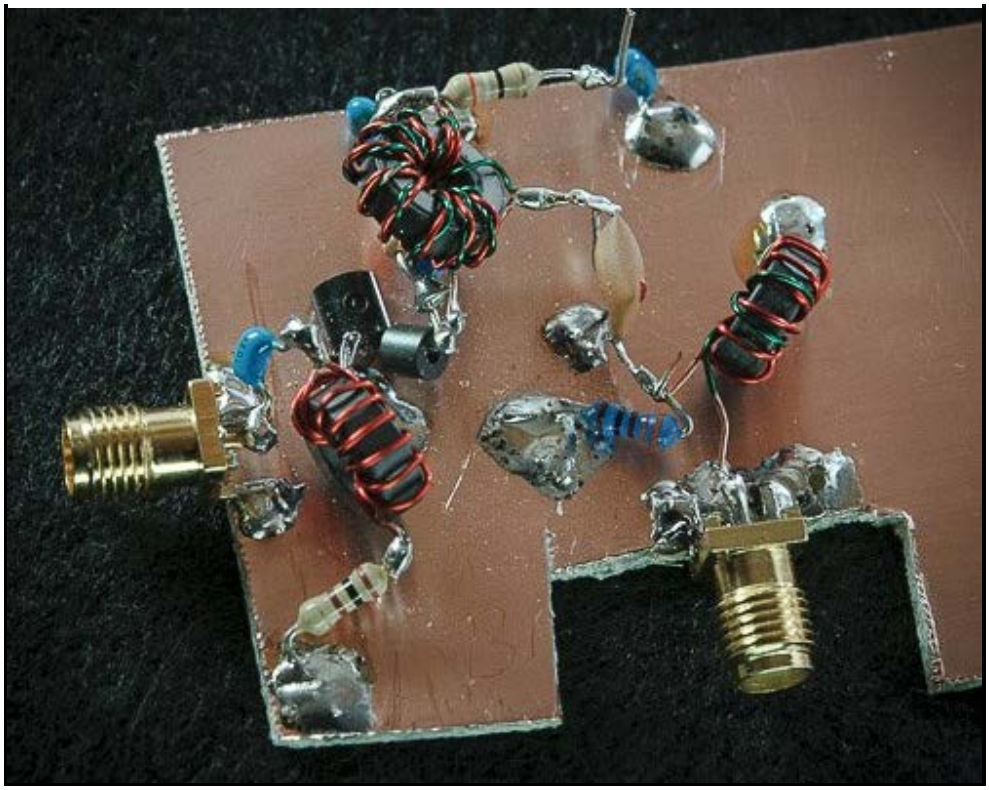




10.7 MHz IF Diplexer

Oct 19, 2013

VE7BPO



Sundry Experiments 2012 - 2013

This page shows some of my better non-VHF experiments for 2012-2013.

Although VHF and UHF excite me greatly, It's always fun to build at HF, or even AF.

Section 1: I explain why you might see sweeps that look like oscilloscope tracings on QRPHB: they're devices swept with equipment designed by Bob, K3NHI. Bob's work simply amazes me — full-on, creative precocity.

Section 2: An LM1875 AF power amplifer test.

Section 3: Three Questions with Jason, NT7S.

Section 4: EMRFD Experiments — A 1-on-1 Tracking or Offset Phase-locked Loop.

Section 5: Boot-strapped Popcorn AF Feedback Pair.

Section 6: Non-Mechanical Iambic Paddle.

Section 7: A Journey Above HF.

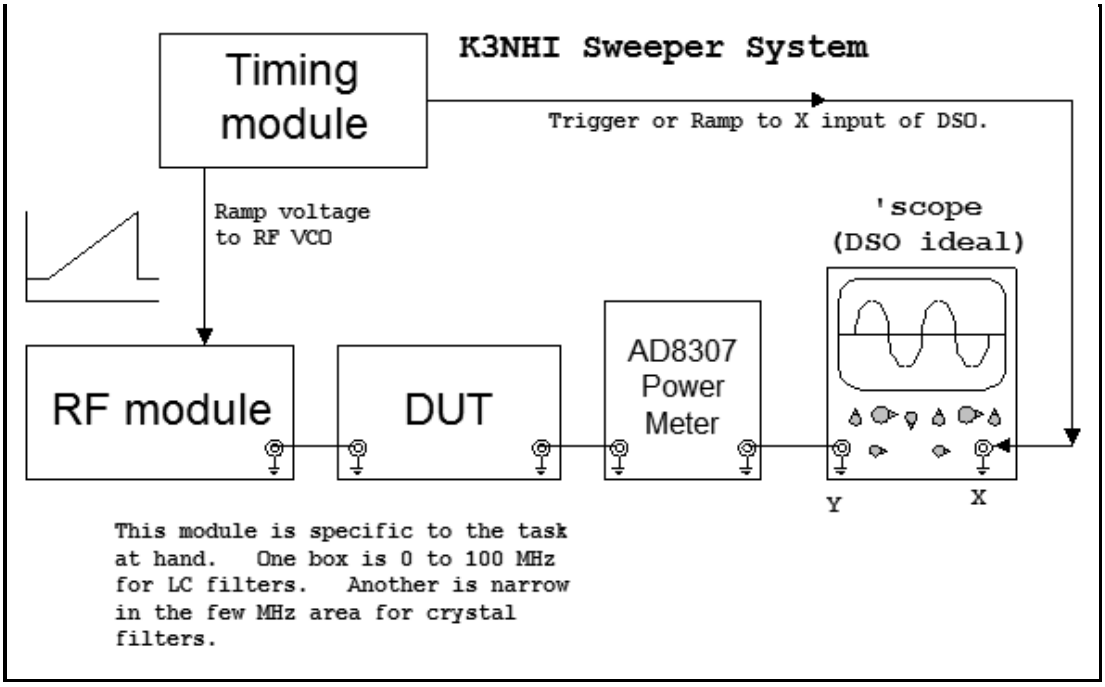
Section 8: Popcorn AF Amplifier — Reprise

Section 9: The Progressive Receiver by John, K5IRK and Wes, W7ZOI

Section 10: Miscellaneous Pictures and Figures

1. Analog Sweep System

Today, advanced experimenters might build a network analyzer/sweeper incorporating a microcontroller, a DDS, or Si570 based frequency synthesizer, plus the needed analog RF circuitry. I went another direction: the Bob Kopski, K3NHI sweep system — all analog, no lines of code and probably 4X the bench work. I show some photos, traces and text in hope it might inspire you to pursue your own sweep system — digital-based or otherwise.



Above — The K3NHI Sweep System macro diagram drawn by Wes. The resultant trace looks like the output of a tracking generator plus spectrum analyzer. What I like most is that I'm measuring with my "tough" 'scope and need not worry about input power and so forth like we do with expensive RF test gear. I simply love measuring signals with my oscilloscope. Testing circuits with Bob's sweep system compels me to treasure component-level analog design and renews my passion afresh.

Above — 3 components of my K3NHI sweep system. Click for a higher resolution photo. To date, I've made the Utility Sweep Generator (time base) *on the bottom*, a 1-118 MHz VCO with clean and level output *top left*, an AD8307 power meter optimized for sweeps *top right* and a crystal filter VCO *shown later*.

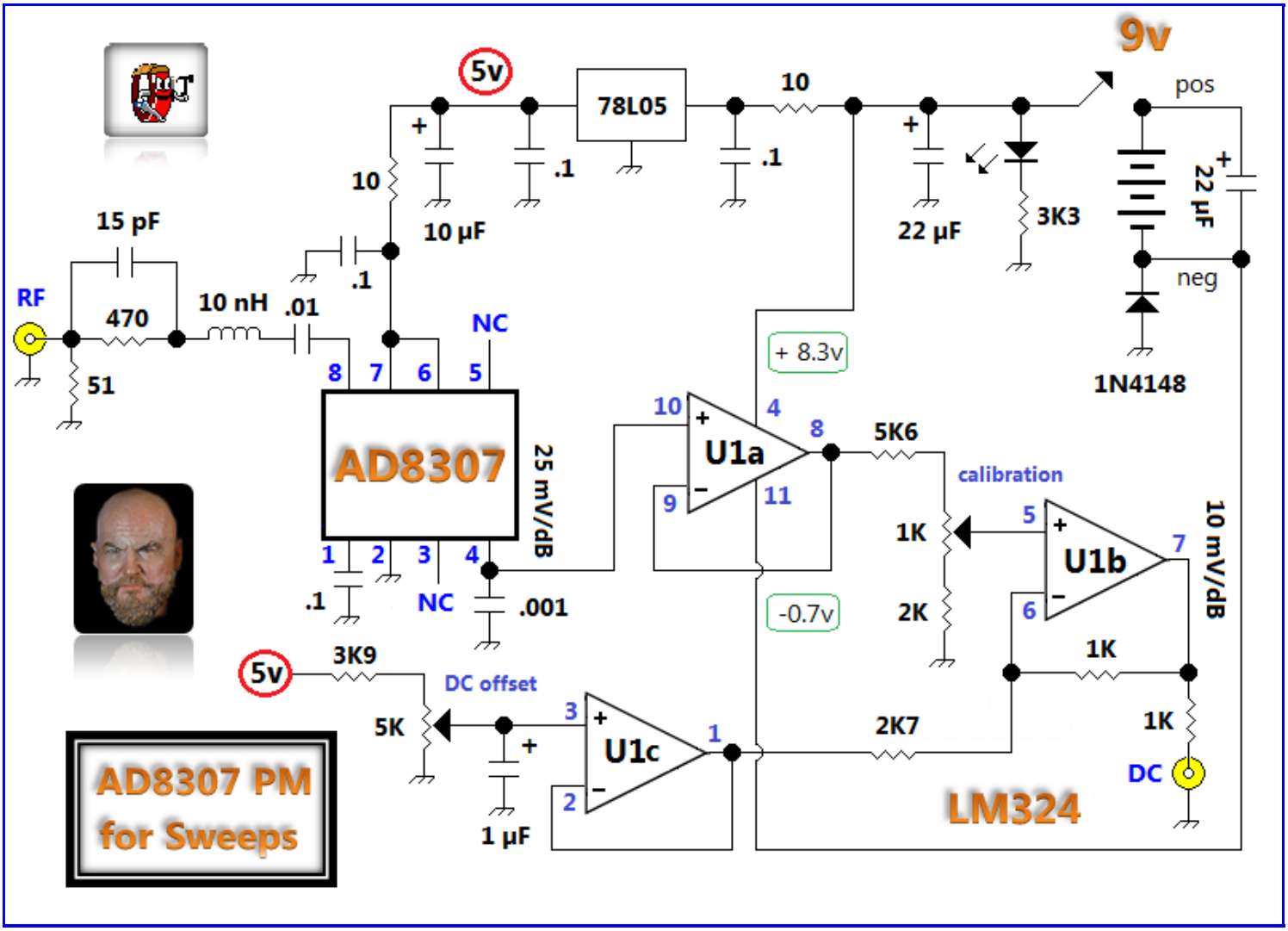
Building Bob's Utility Sweep Generator proved difficult. Although technically just a ramp generator, this 1 is calibrated, provides high isolation between the X and Y channels and will sweep *anything*. The power supply has 8 different regulated DC voltages including ~ - 3V. I may use it as the time base for a spectrum analyzer project 1 day.
Bob gave me permission to post the schematics: One Two Three IC1-5 is an LM324 op-amp. Doc 1 Doc 2

Email me for some build notes.

Click, click, or click for photos of the 1- 118 MHz VCO. Click for the schematics. One Two

I show a bare-bones AD8307-based Power Meter (PM) on the **RF Workbench 5** web page, however, to augment the PM for sweeping, I added input compensation, plus some tweaks from Bob's QEX articles: Bob Kopski, K3NHI — An Advanced VHF Wattmeter, QEX, May/June 2002 *and* Bob Kopski, K3NHI — A Simple Enhancement for the Advanced VHF Wattmeter, QEX, Sept/Oct 2003.

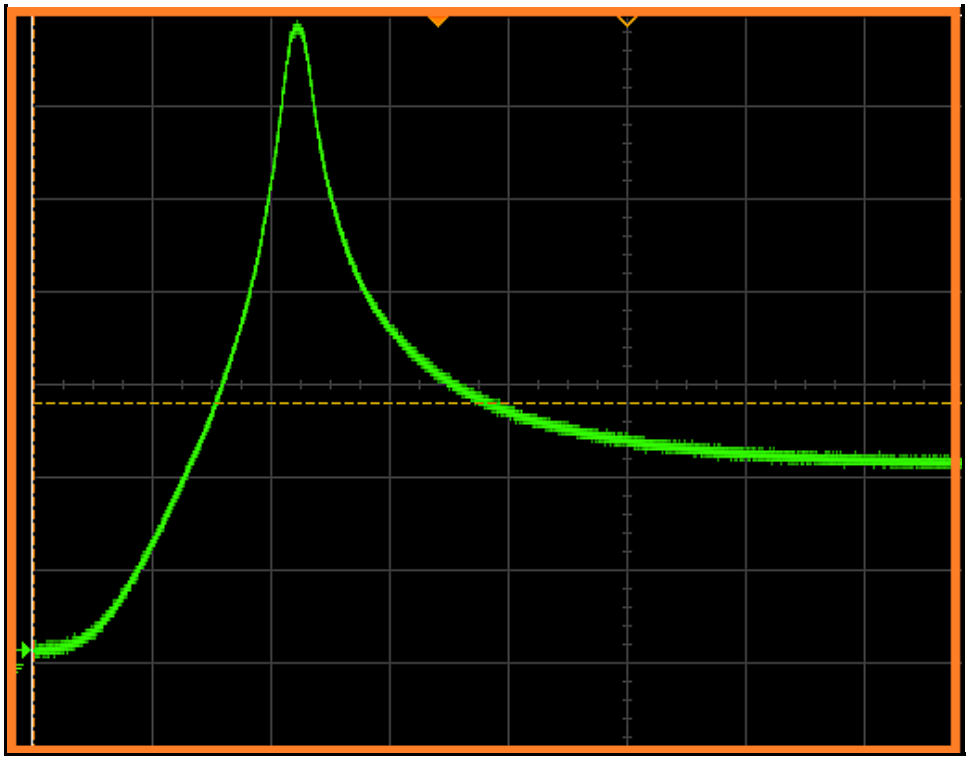
I strongly recommend you build Bob's power meter (referenced above) if you're contemplating a power meter build.



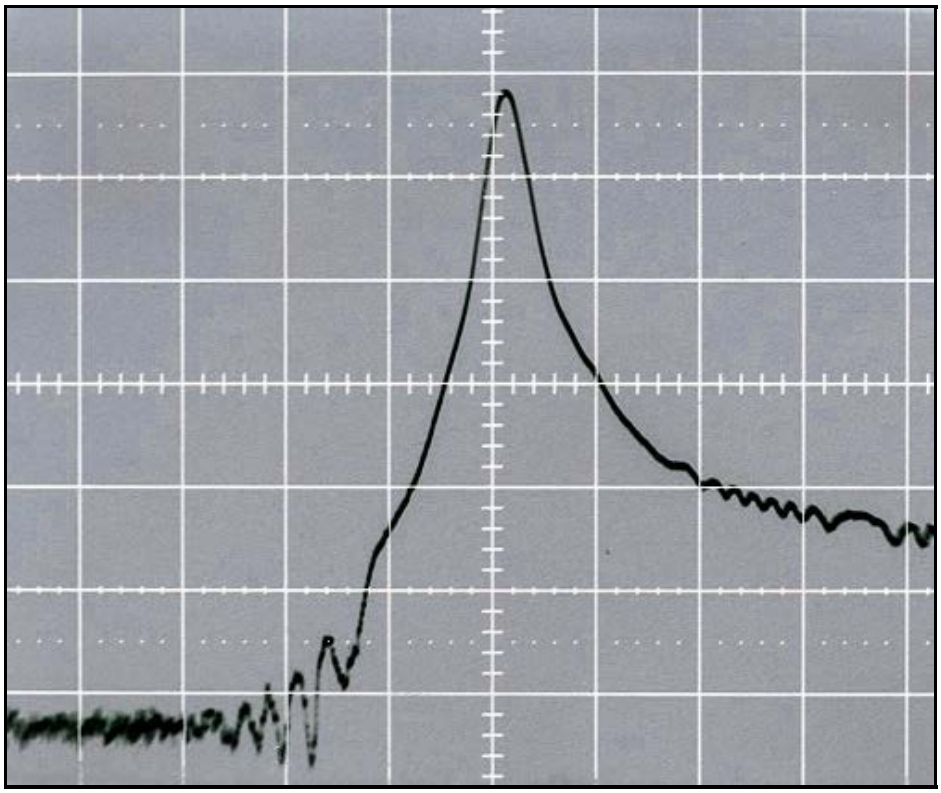
Above — My AD8307-based power meter. Click for a photo of an early version lacking the level shifter. The level shifter, or DC offset control allows precise Y axis control to enable a resolution up to 1mV/dB when set up properly in the DSO. With this resolution, it's possible to see filter ripple.

Above — The K3NHI sweep system in action. This photo shows the power meter with the (later) added DC offset control. Using the offset control potentiometer, I'm able to examine the top of a signal peak at 1 dB per division with various spans.

Above — I swept a 7 MHz band-pass filter bench module.



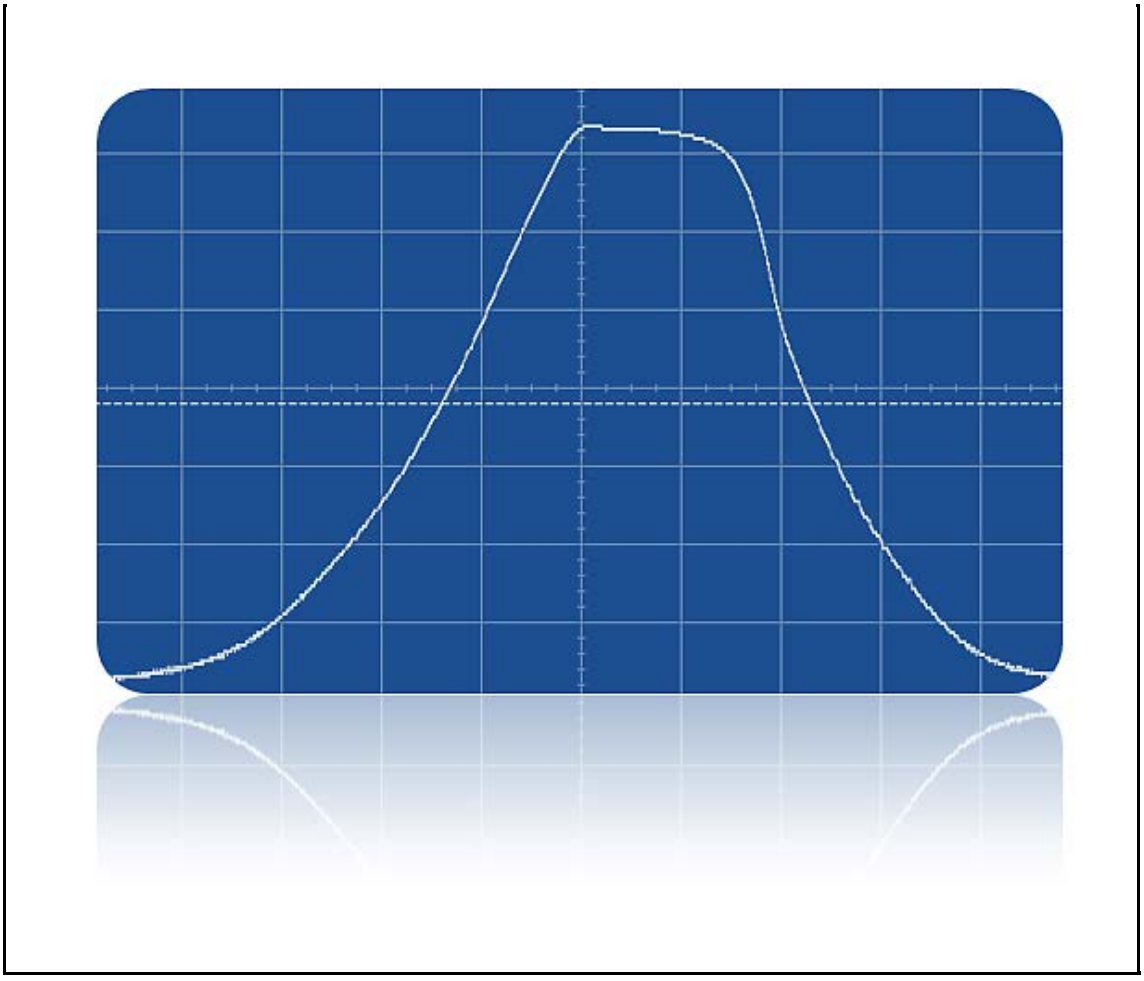
Above — A sweep of the above 7 MHz band-pass filter. At the time I was still learning system calibration and remember feeling blissful that I made such a cool sweep system.



Above — A sweep of the 7 MHz band-pass filter with a tracking generator plus spectrum analyzer for comparison. The span differed from that used with the K3NHI sweep system, however, I'm sure I'll get better with the K3NHI system over time.

Above — My build of the K3NHI *Hartley* VCO for sweeping crystal filters. I ordered a 4 mm tuning knob for the 10-turn "Manual Tune" potentiometer used to help center the sweep in my 'scope. This 10-turn pot, a DIP switch with C's and a secondary L, plus two MV209 varactors allow narrow resolution sweeps (< 5 Hz nominal) within a ~2.5 - 18 MHz range.

I monitor the sweep frequency on a counter via the VCO monitor port and all the tweaks on this VCO and the Utility Sweep Generator allow easy filter centering. Click for a breadboard of the VCO with the first buffer and a temporary BNC connector for testing. Click for the whole project. The secondary wideband buffer provides strong signal fidelity, reverse isolation and output return loss (22.3 dB) — it draws 61 mA. Click for the VCO schematic courtesy of Bob, K3NHI. Click for a side-by-side of a xtal filter as measured with Bob's sweep system plus an N2PK VNA.



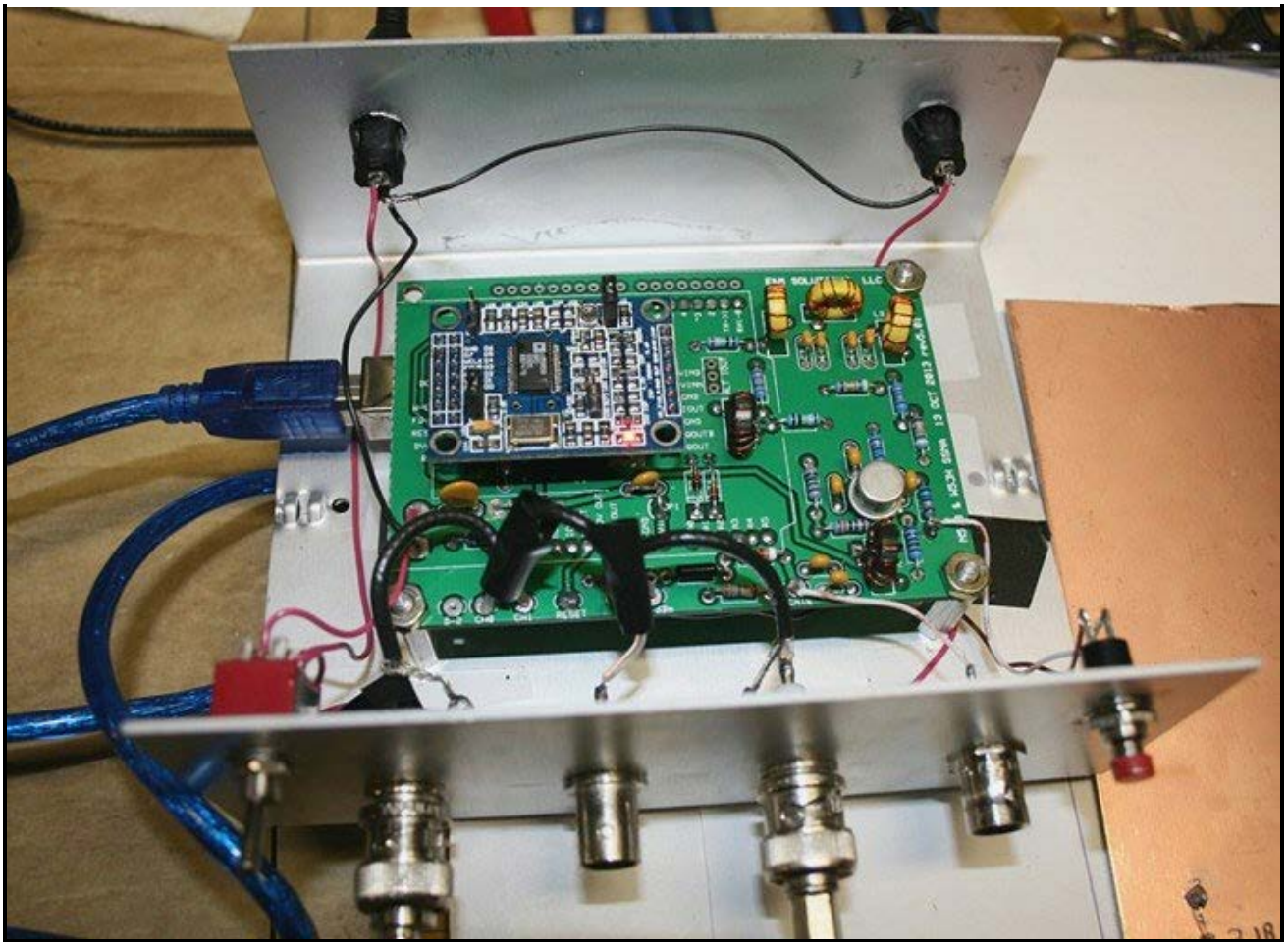
Above — I made and then swept a simple ~500 Hertz wide 4.9152 MHz Cohn or Min-Loss filter using 4 crystals.

So now, if you see sweeps on the site that look like oscilloscope tracings, you'll understand how they were created. I've learned so much from Bob's work and his mentorship last Spring. John, K5IRK coached me also.

QRP-POSDATA for October 2013

3 builders incorporated Bob's sweep system circuitry into projects including a receiver, a spectrum analyzer and the following build of the Utility Sweep Generator (USG) by Jay: Jay built some PC boards to simplify the wiring challenge this USG presents. Click 1 Click 2 Click 3. Great stuff — thanks for sharing.

QRP-POSDATA for March 2014 — Poor Hams Scalar Network Analyzer (PHSNA) —



Above — The PHSNA built by Mikey, WB8ICN.

Jerry W5JH, along with Jim, N5IB and Nick, WA5BDU developed this low-cost sweeper/ lab toolkit.

The Poor Hams Scalar Network Analyzer consists of an Arduino UNO R3, plus an AD9850, or AD9851 DDS, a W7ZOI/W7PUA Power Meter and a MS Windows OS based computer. Builders can scan and plot L-C filters, crystal filters, RF amplifiers and such — much like Bob's sweep system — or an HF tracking generator + spectrum analyzer. The PHSNA also measures crystal parameters with little fuss. Connect a return loss bridge to easily sweep return loss measures of the input or output ports of filters, amplifiers, antennas and more. The total cost to build the PHSNA is approximately $50-60 USD.

Mikey graciously sent me some photographs of his PHSNA build. Complete system in his lab with the chassis lids removed for these photos. Monitor photo showing menu choices. Power meter. Mikey's jig to examine crystals with a 12.5 Ω termination. Once you own a calibrated sweep system, you'll wonder how you ever managed without 1.

Thanks again to Mikey for the photos. I built and tested the return loss bridge using the PCB from the PHSNA Yahoo group. Click for a 613 KB pdf file of my build. Nick, WA5BDU web site.

Hats off to Jerry and crew for this open-system project! A Yahoo group called PHSNA serves as the communications hub and houses superb, detailed documentation. You need to join Yahoo to access this group. Then search for PHNSA and while your at it, also sign onto the EMRFD group.

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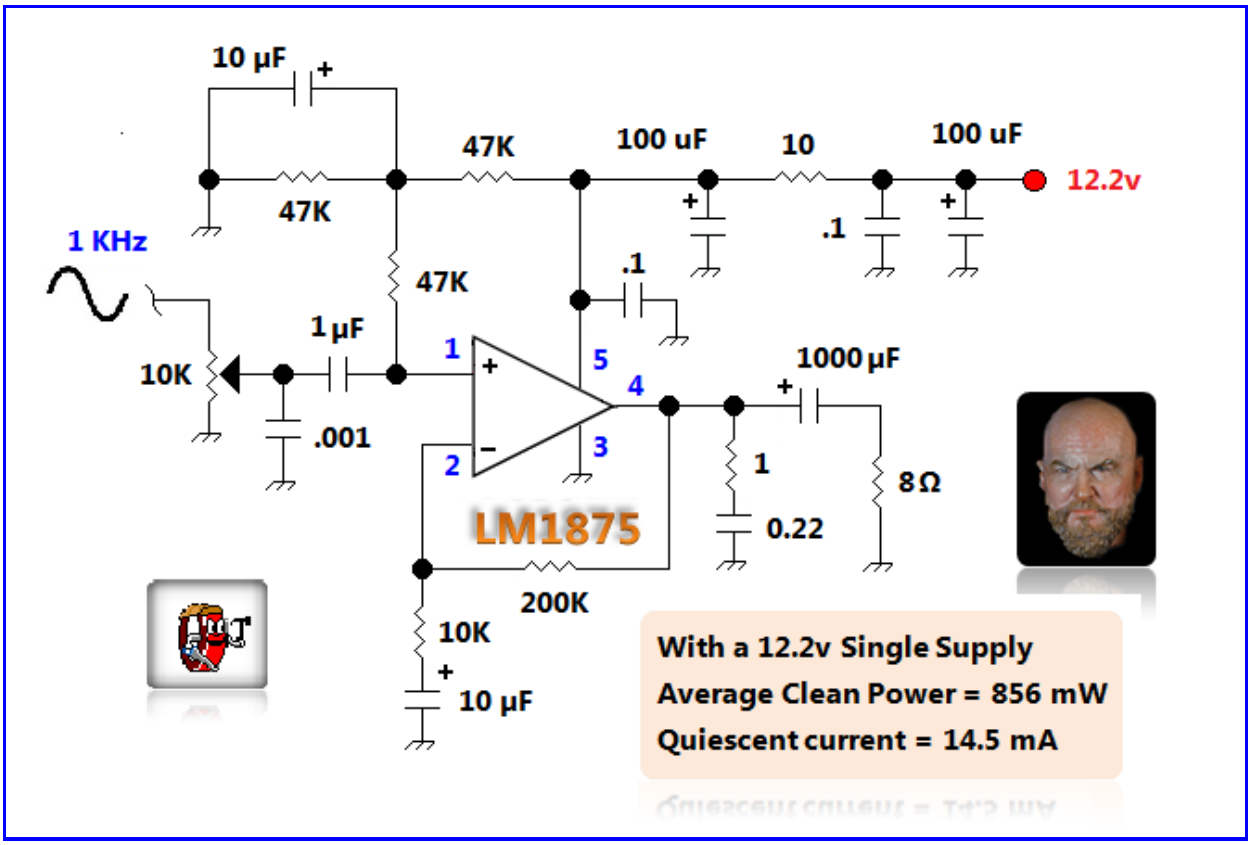
2. LM1875 Audio Power Amplifier

I tested the LM1875 AF power amplifer because its specifications look great: 20 watts into a 4Ω or 8Ω load on ±25V supplies and a TO-220 package for easy heat sinking. Of course, for this web site, I tested it with a typical radio experimenter bench power supply; a single-supply at ~ 12 VDC.

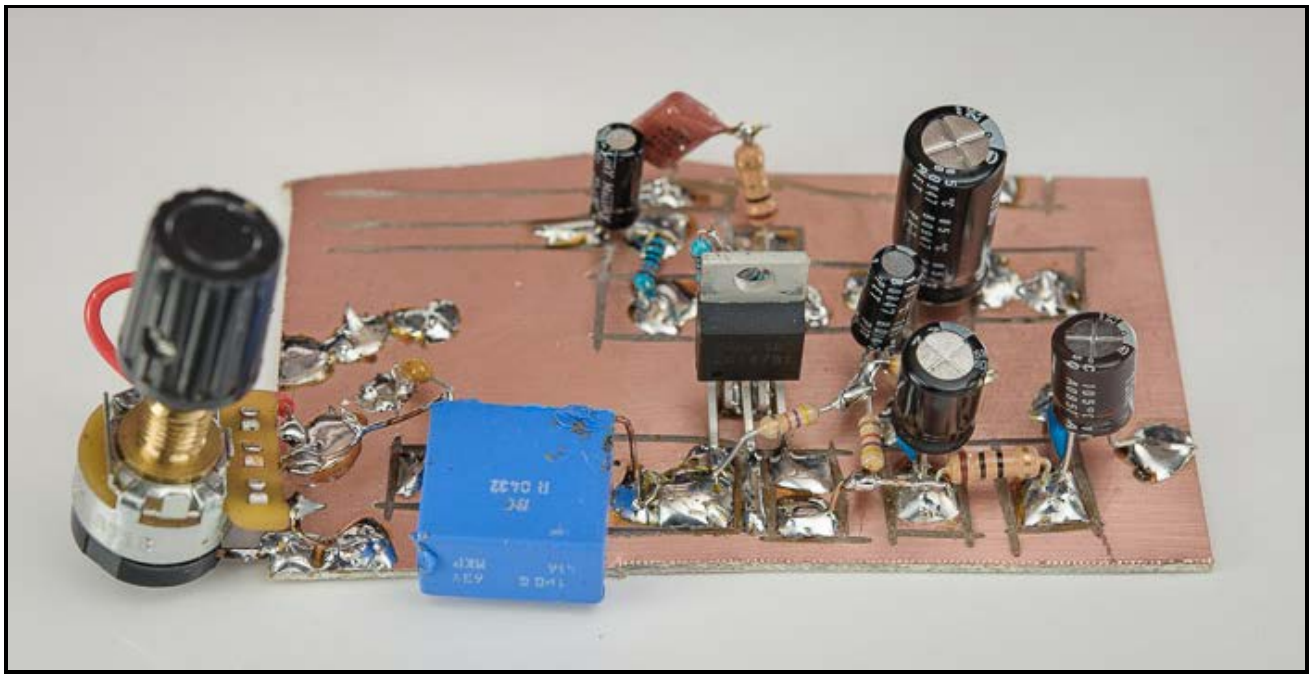
This is probably not a great part for Ugly Construction and I *attempted* to return the load ground, the output Zobel RC filter network, feedback loop and input grounds to a central grounding point through separate paths cut paths into my copper board. A better breadboard method might include the so called "star grounding".

I saw RF oscillations on the 'scope and removed them by soldering a 0.001µF bypass capacitor across the input. A 470 pF bypass capacitor did not work well enough. The datasheet describes specific causes and cures for RF oscillations and I've learned they must be heeded. I once found

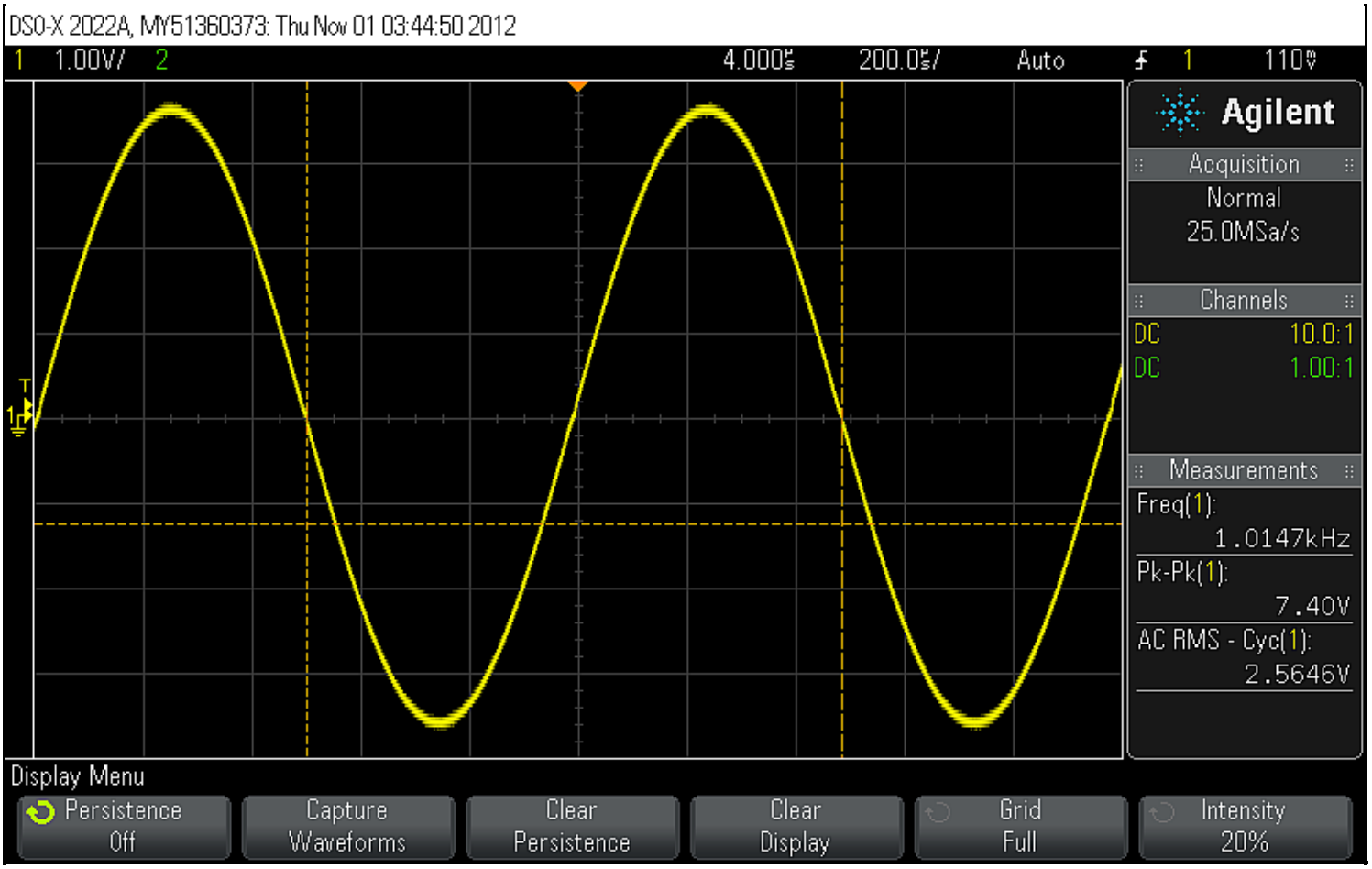
similar problems with an LM380.



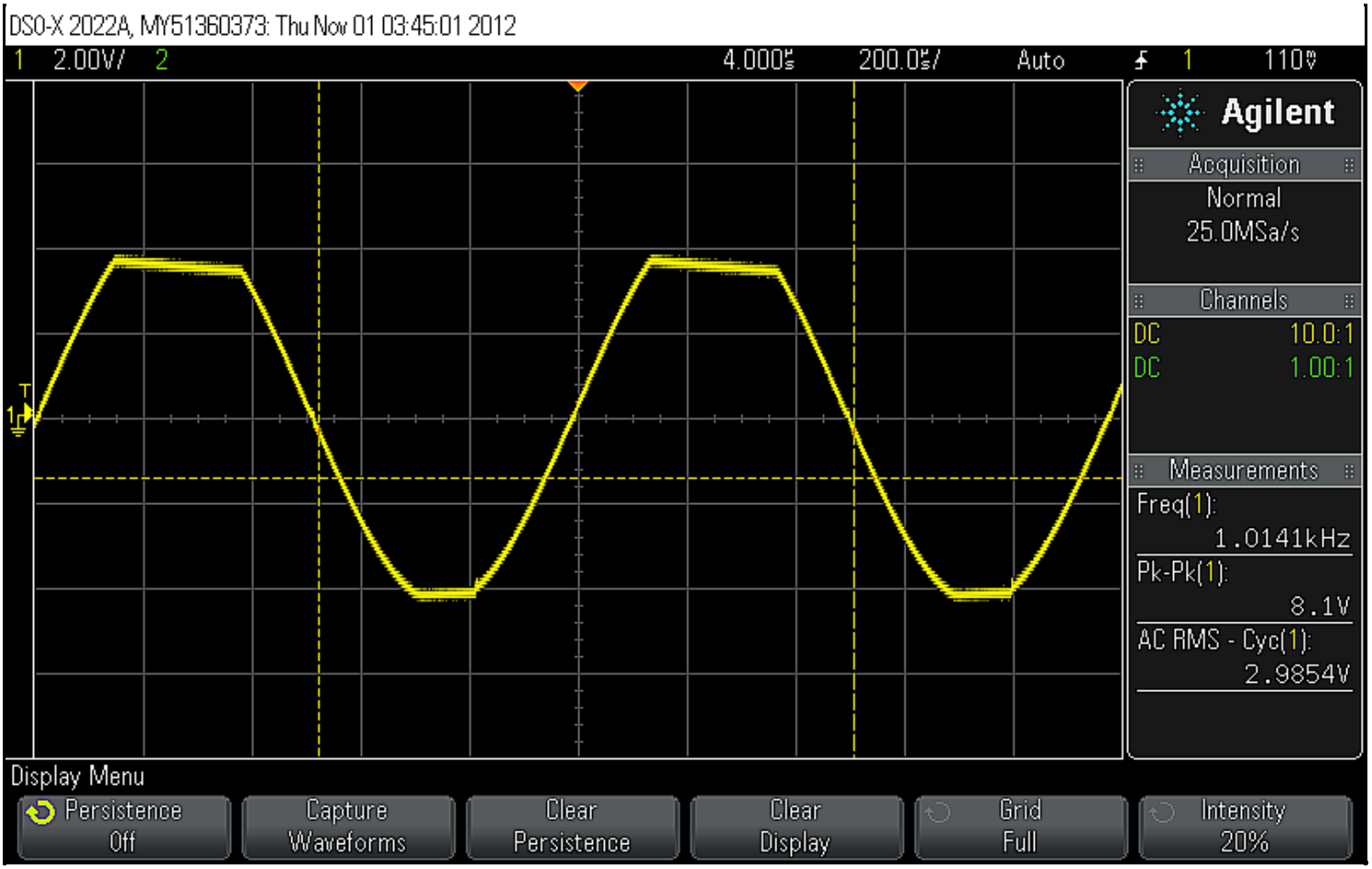
Above — The LM1875 in the ~suggested datasheet, single-supply set up. This amplifer reminds me of setting up an op-amp. Unlike the LM380, within limits, you may choose the gain to suit your needs. As shown the gain = 25.6 dB. Dropping the feedback R to 100K dropped the gain down to 20.5 dB. For clean output power capacity; it blows away the LM380.



Above — My first LM1875 test breadboard. After the photo, I moved the 0.1 µF RF bypass cap right onto lead 5 — we should carefully RF bypass device power supply leads, but I got sloppy.



Above — At the maximum power before visible sine wave distortion appears; 856 mW. I listened to this amp while connected to a line-level tape player and an 8 Ω speaker load: very nice. I want to try it with a split +/-15 VDC supply and a star grounded breadboard since a 12v single-supply limits the output power so much. Still, at 12 volts single-supply, this IC yielded the highest clean, average output power of any AF Power chip I've tested: a worthy consideration for a high-grade receiver.



Above — 'Scope tracing; I advanced the 10K volume pot to drive the amp into clipping.

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3. Interview — Three Questions with Jason, NT7S from Etherkit

I follow 2 English language amateur radio blogs — 1 is Ripples in the Ether by Jason, NT7S

Possessing a modern flare, Jason, the blogger, gently but resolutely challenges some of the cliquish, dogmatic thought and behavior that tarnishes amateur radio, or just blogs about fun stuff. He writes well — creating an emotional dialog that stimulates thought and reflection. We get a sense that he cares about our radio hobby and wants it to grow and improve.

Jason, the man behind Etherkit, champions a modern, open-source vision that I find both positive and refreshing.

1. Tell me about your decision to embrace the open source software philosophy for your hardware in a time where proprietary code, copyrights and patents still hold strong. How do Ham Radio equipment sellers benefit from code sharing?

I believe that the open sharing of knowledge has always been one of the cornerstones of our amateur radio community, going back to its earliest days. So the open source/open hardware ethos has always resonated with me in regard to our hobby. I started Etherkit with the intention of providing a small bit of income to my family and as a way to promote the idea of open hardware within the ham radio community at large. I have no illusions of becoming the next Elecraft, but I hope that I can build up a stable of affordable and fully-open ham radio kits that will be "hackable" and extensible for the motivated experimenter. I do this by providing the full source code for my microcontroller firmware, all of the PCB design files, Creative Commons licensed documentation, and programming ports for my products. I've already seen some neat examples of

customers extending my first product (the OpenBeacon MEPT kit) by doing things such as adding in WWVB time discipline and pairing it up with a Raspberry Pi for cheap automation. I hope that others will take my code or my circuits and re-purpose them in their own work, even if they don't buy my products.

I am not an open source zealot and do not begrudge the large majority of vendors who choose to keep their intellectual property closed. However, most of what us smaller companies do is not on the cutting edge of radio. We leverage the knowledge and works of those who came before us. Perhaps if I created something wholly-new that would be patent worthy, I would consider keeping it closed, but that's not the kind of products that I'm able to develop as a one-man operation. We do not copy the designs of others, but we do take concepts that are for the most part well-tested and come with new ways to implement them. Because of that, it's my personal opinion that I have a duty to keep my designs open.

In the open hardware world at-large, there is a discussion about whether open source hurts your own business prospects. There are still some debatable points in that discussion, but I think it has been shown that if you look at the entire balance, open hardware is a good thing for smaller companies. One of the largest concerns is that under most open source licenses, a competitor can just clone your hardware and undercut your sales. That is a genuine concern, but I think that products such as Arduino have shown that if you make a quality product, most folks will recognize that quality and stick with the original.

Even if others buy a clone of your hardware, in all likelihood, that may be strengthening your brand identity (as long as that vendor isn't stealing your name). Another concern is that a customer can just copy your product for themselves. To that, I say good!. Because of the work and costs involved (economy of scale), it's going to be time and/or money consuming to make that copy. It's probably cheaper and faster to just buy the kit. The reason you copy it for personal use is because you love working with the technology. Which is exactly what I want to encourage. You may lose a small bit of sales, but I think it gives you more name recognition in the end.

2. What's it like being a vendor at Dayton?

To be clear, I wasn't a vendor at Hamvention in 2012 (hopefully I will be there by 2014), but I was a vendor at Four Days In May at the Fairborn Holiday Inn. It was a wonderful experience to get to sit with the big names in the QRP world, selling my wares. I got the opportunity to meet tons of QRPers and build up some good relationships. Online sales are wonderful for the ultra-small operations such as myself, but nothing beats actually meeting your customers face-to-face, especially when you are at the world's most well-known QRP convention.

3. In industry, SMT parts are normal and hole-thru might better be called "hold-over"; what's your view on kitting products with SMT parts?

We've seen some SMT kits within the QRP world, but they still are more of an oddity than anything else. I understand the concerns that people have with SMT assembly, but I think that there is still a lot of trepidation that builders needn't have. It's my opinion that SMT construction with "larger" components such 1206 or 0805 is well within the capabilities of the average kit builder. I also believe that once you are comfortable with SMT construction, it is probably faster and more efficient than through-hole construction.

OpenBeacon is a through-hole product, but I have had a QRP CW rig in development for the last two years that is a SMT design. In beta testing, I've found that one of the biggest challenges in kitting is that I have to clearly identify each and every component. With a through-hole kit, you can just throw all of the resistors or all of the capacitors together because they are clearly marked. Not so with SMT. You have to have a system to keep each value separated from the others and marked with a value. SMT resistors and semiconductors have a laser-etched value, but it's nearly impossible to see by naked-eye, and SMT capacitors generally have no markings at all.

So I have had to compartmentalize each strip of components of the same value cut from a reel, and mark them with a sticker. That is pretty costly and time-consuming. I'm hoping to find ways to streamline this process so that I can release SMT kits without the large time investment that it currently takes.

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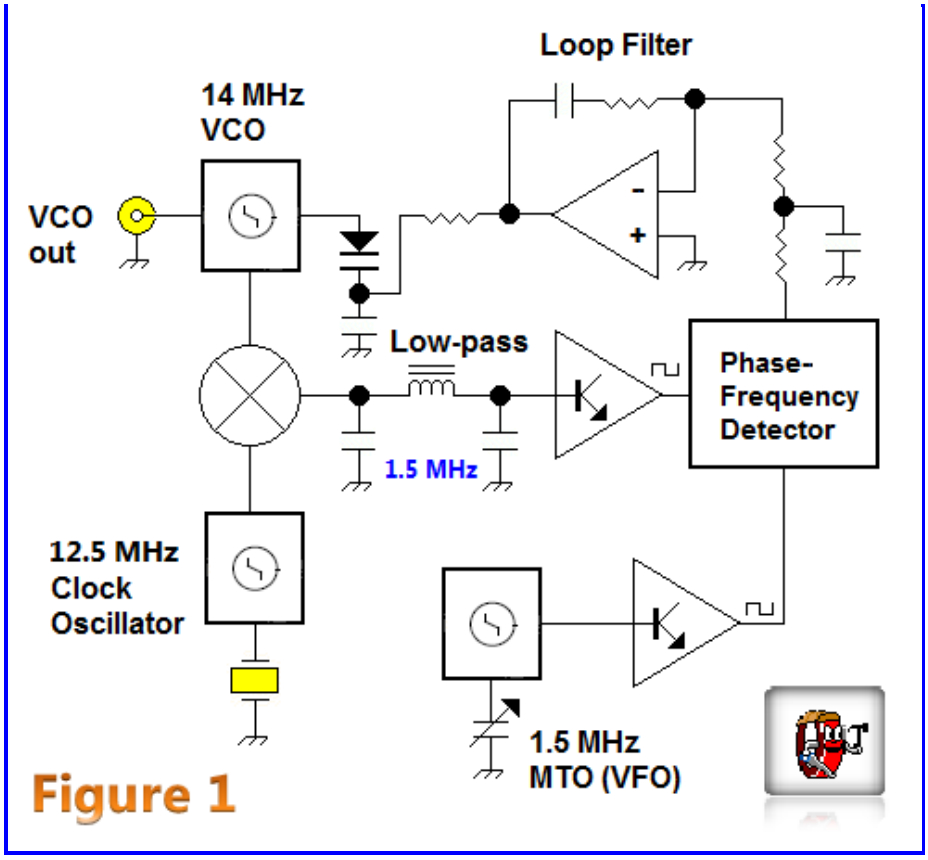
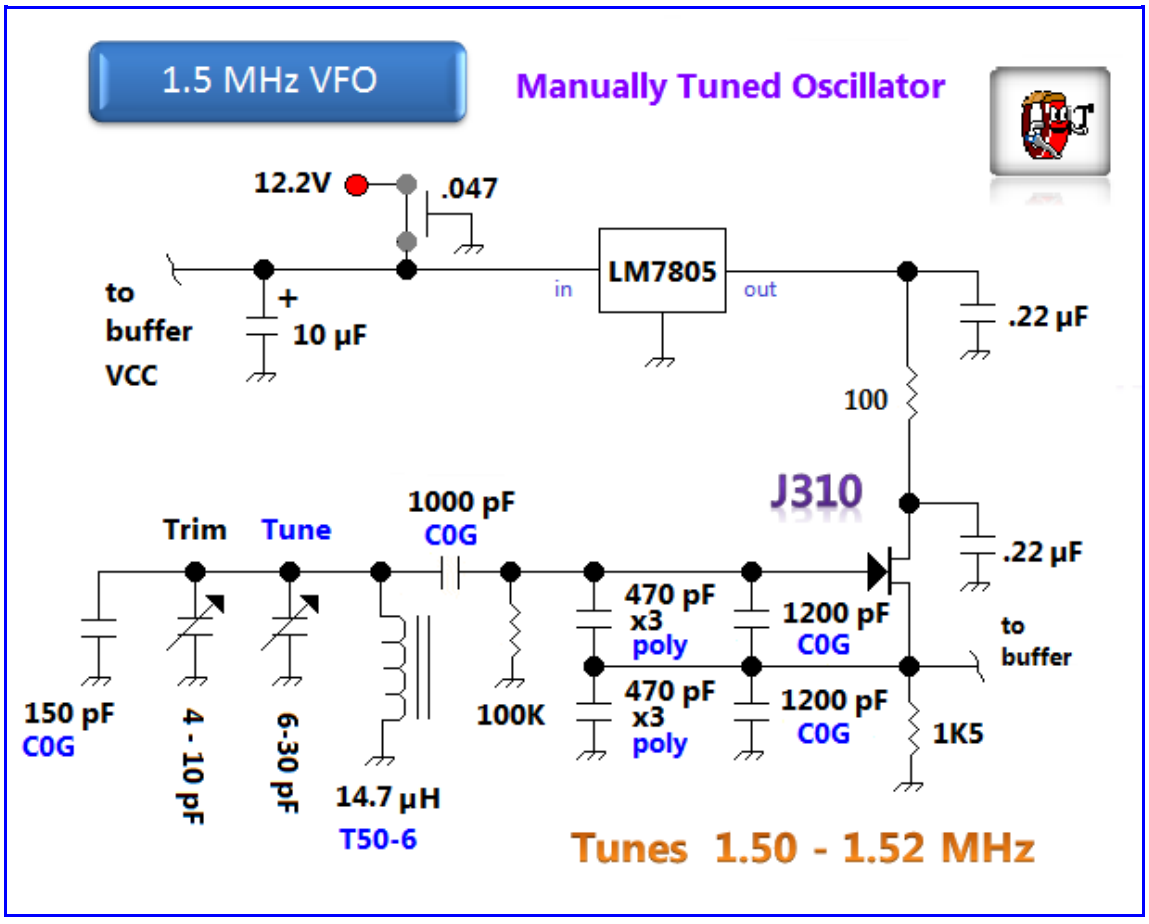
4. EMRFD Experiments — A 1-on-1 Tracking Phase-locked Loop

I built the 1-on-1, or offset phase-locked loop circuitry described on EMRFD page 4.22 and share these schematics in faith you'll create your own. Rich with wisdom and reason, this section lies among the best topics from EMRFD. Please read Wes' notes since I won't repeat his narrative — only supply a few ideas and measurements. In the article closer, Wes suggests some modern parts to raise performance and I applied all of them with the exception of the 14 MHz VCO.

Rather than building the main VCO with divide by N circuitry to allow multiband use, I copied the original 14 MHz oscillator verbatim. Why? Well, I wanted to test this VCO: a design that wisely doesn't expose the varactor to high impedance or signal amplitude and thus avoids forward-biasing the single tuning diode. I've discussed this before on the QRP Modules 2011 web page under **7 MHz VCO Experiments**. Also, I really just wanted to learn about PLL circuitry. The Figure 1 macro schematic below illustrates this project.

In my circuit, a frequency stable 14 MHz VCO = **the goal**; the rest of the circuitry supports this.



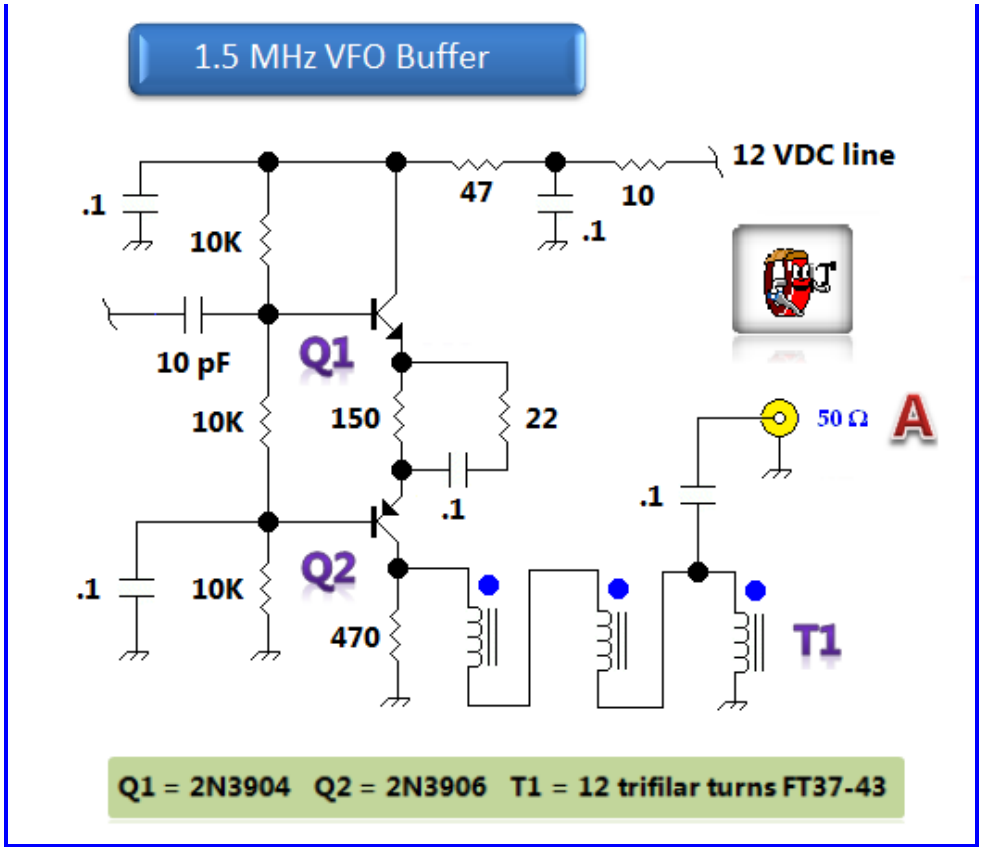


Above — The 1.5 MHz VFO. In his modern writing, Wes calls this the MTO, or Manually Tuned Oscillator *in the context of a tracking PLL.* I wound the L with # 30 AWG wire on a T50-6 toroid.

My MTO exhibits a low tuning range (only 1.50 to 1.52 MHz) since I built in a box with a small air-variable capacitor that swung only 24 pF and I ran the "Colpitts capacitors" at 2610 pF to keep phase noise low. This box normally holds a VHF oscillator and I just removed the main board and swapped in a 1.5 MHz equipped copper board. I won't keep this PLL and thus sticking the 1.5 MHz MTO in an existing oscillator chassis with a grounded tuning shaft and feedthrough capacitor helped save money and time.

Above — The built 1.5 MHz MTO. With temperature compensation from 6 stiff-leaded, 600 VDC, 470 pF polystyrene capacitors, my frequency drift measured between 3 and 4 hertz per hour upward at room temperature. Properly designed + built + temperature compensated L-C oscillators at 1.5 to 3.5 MHz may exhibit stellar temperature stability. See the VFO - 2011 web page for some tips.

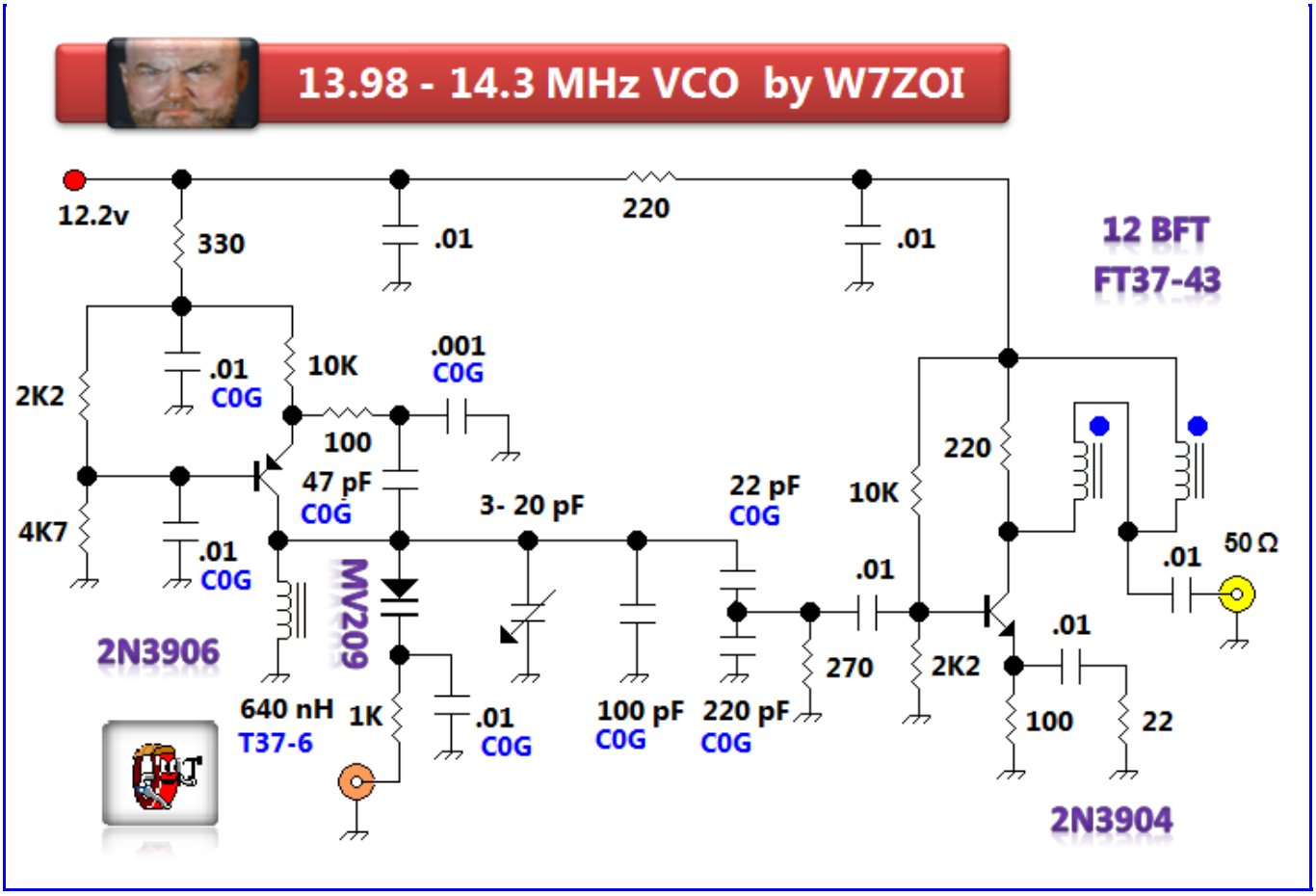
Since this VFO was sublimely frequency stable, I didn't possess the guts to change up the L-C ratio to garner a wider tuning range from the small air-variable tuning capacitor. A 100 pF, or greater delta-F air-variable tuning capacitor would stretch the VFO (MTO) tuning range nicely.



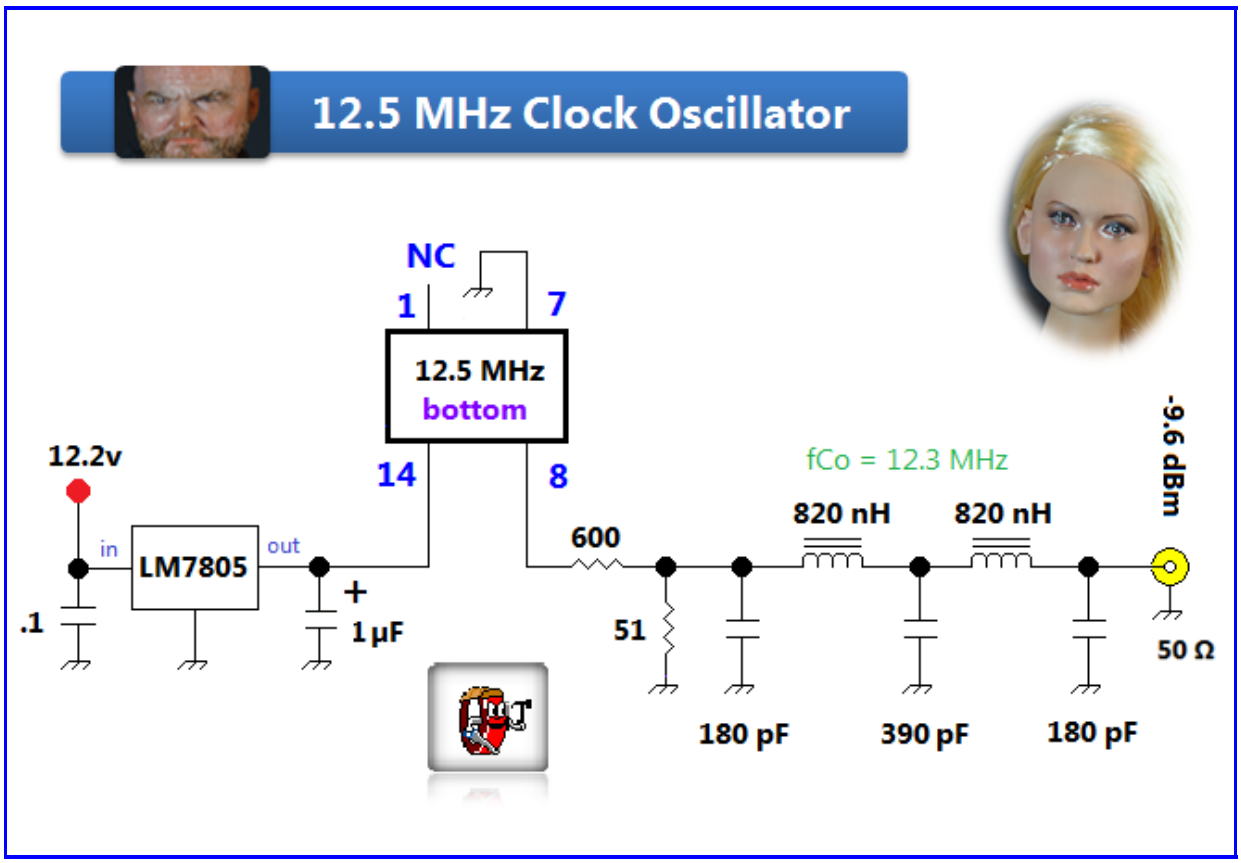
Above — I designed this buffer last year and it's my new favorite. Click for the original. A 10 pF C0G/NP0 capacitor lightly couples the MTO output to the high impedance of Q1, an emitter follower. Further, a common base amp provides gain and *essential* reverse isolation. You may adjust Q2 gain by changing the degenerative feedback offered by the 22 Ω resistor and 0.1 µF capacitor. For example, decreasing the R to 18 Ω may provide 7 dBm output for a diode ring mixer.

MTO output power = 6.71 dBm.

I transformed the 470 Ω collector resistor impedance to 50 Ω with a transmission line transformer. Even though part of the PLL circuitry involves logic gates, or is at DC; as possible, my circuits employ a 50 Ω input or output impedance to allow measurement with my 50 Ω modules and/or instruments, plus transmission via 50 Ω cables.

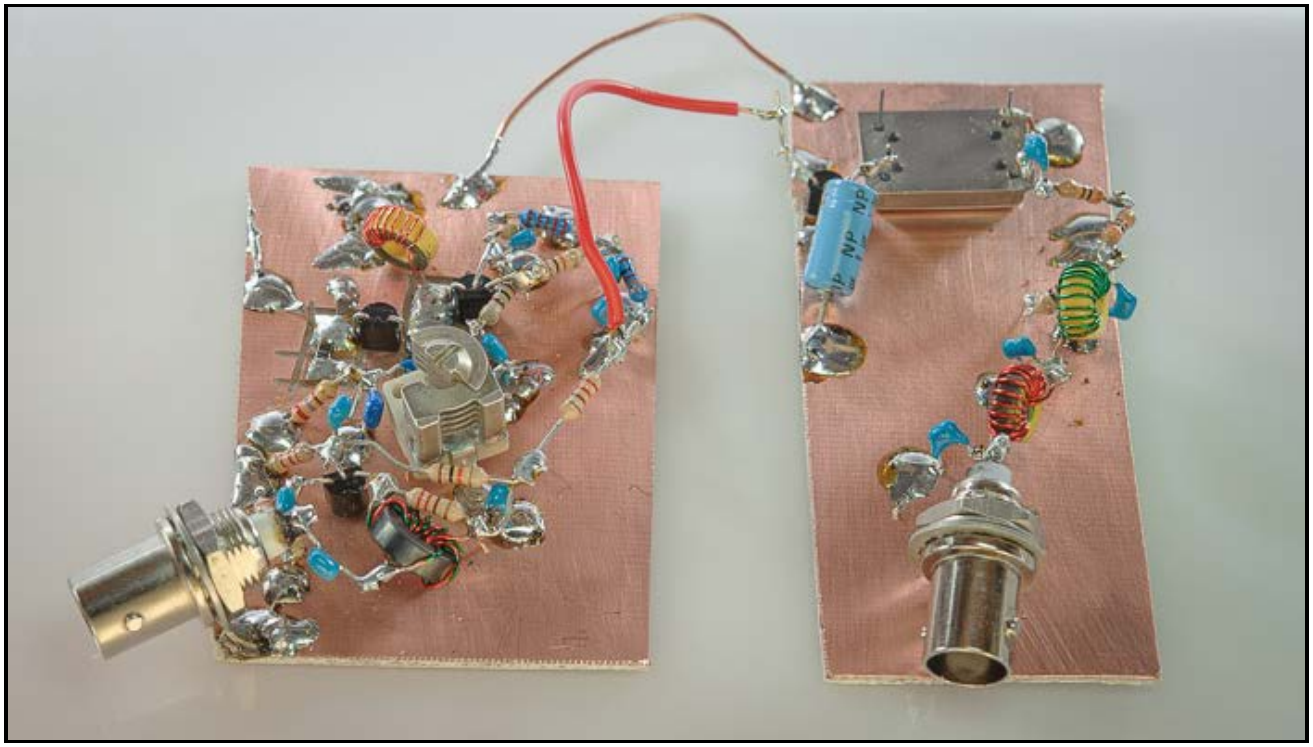


Above — The 13.98 to 14.3 MHz VCO by Wes, W7ZOI (Figure 4.43 in EMRFD). The connector in series with the 1K varactor resistor was an RCA type. Output power = 1.62 dBm. I employed a 3 - 20 pF air variable for the trimmer.

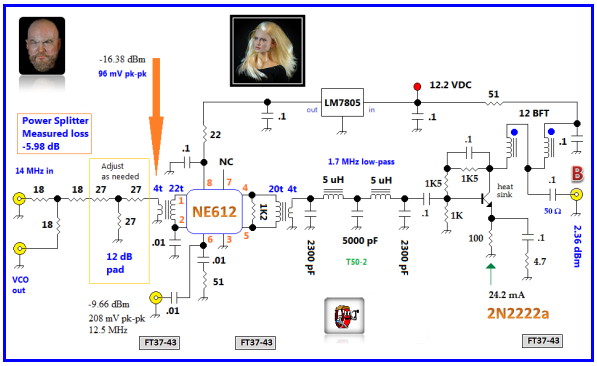


Above — In the original circuit, Wes built his 12.5 MHz crystal oscillator with a single 2N3904. Lacking a 12.5 MHz crystal, I built my xtal oscillator

from an old, junkbox 12.5 MHz clock oscillator. A resistor L- matching network drove a low-pass filter to scrub off harmonics, Click for the clean output 'scope tracing at 211 mV pk-pk in my first version. Later, some tweaks gave a final power of -9.6 dBm (208 mV pk-pk). Many authors switch their NE612 mixers with a peak-peak voltage of ~200-300 mV. An AC-coupled 51 Ω resistor on the NE612 pin 6 properly terminates the oscillator to establish the desired drive power and filtering.



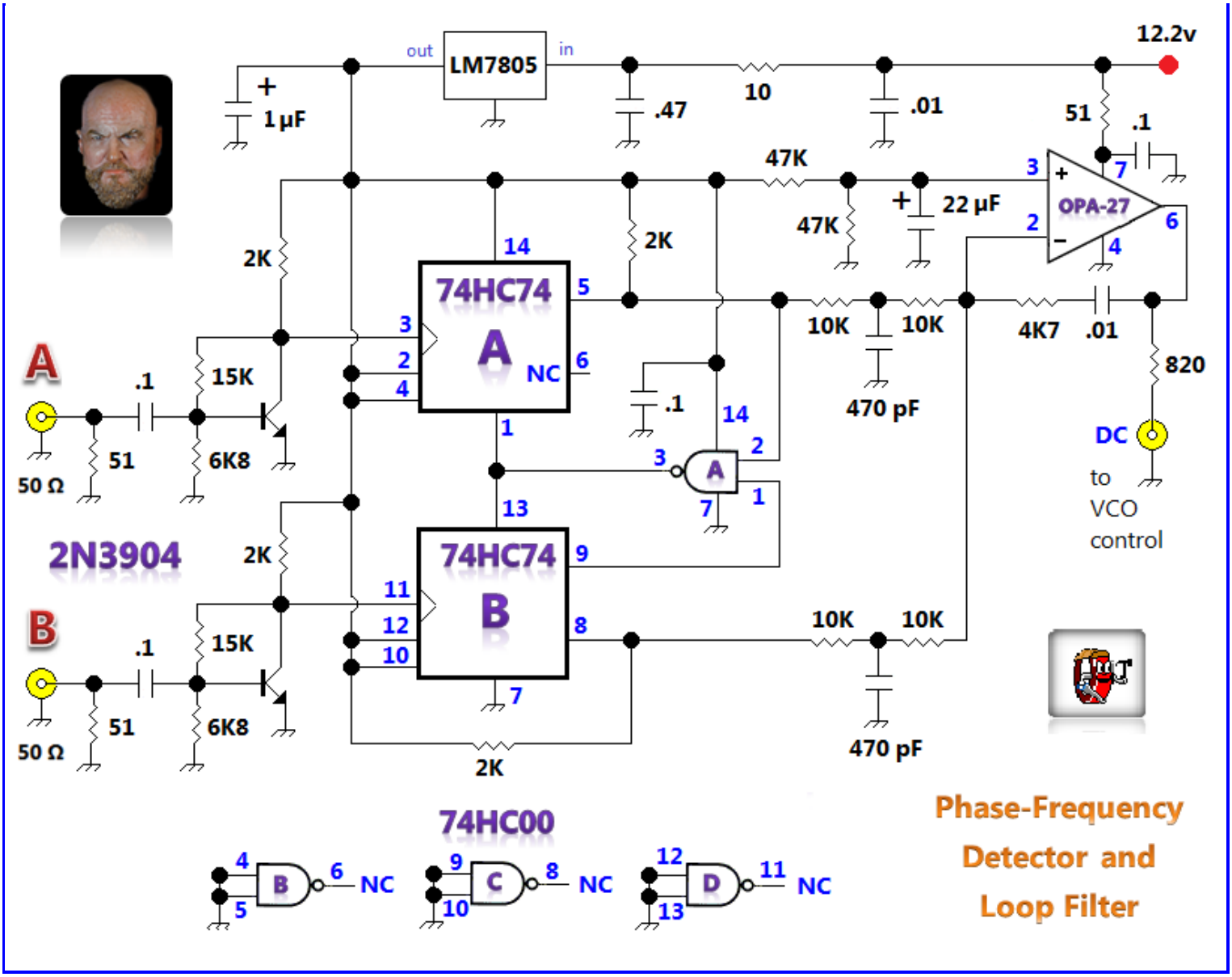
Above — Oscillator breadboards: 14 MHz VCO (left) and the 12.5 MHz clock oscillator (right).



Above — Click to view the power splitter, mixer, low-pass filter and amplifier schematic. Click for an FFT of the clean 1.5 MHz output sine wave. At this point, the 14 MHz VCO has no DC voltage connected to its frequency compensation varactor. As shown, the mixer products are seriously attenuated by the simple, low-pass filter + keeping the mixer RF port signal amplitude low. The power splitter provides the input for the mixer and also the main output for the 14 MHz VCO. The main VCO output requires 50 Ω buffer/amplifer(s) to drive a receiver mixer, transmitter chain, or whatever.

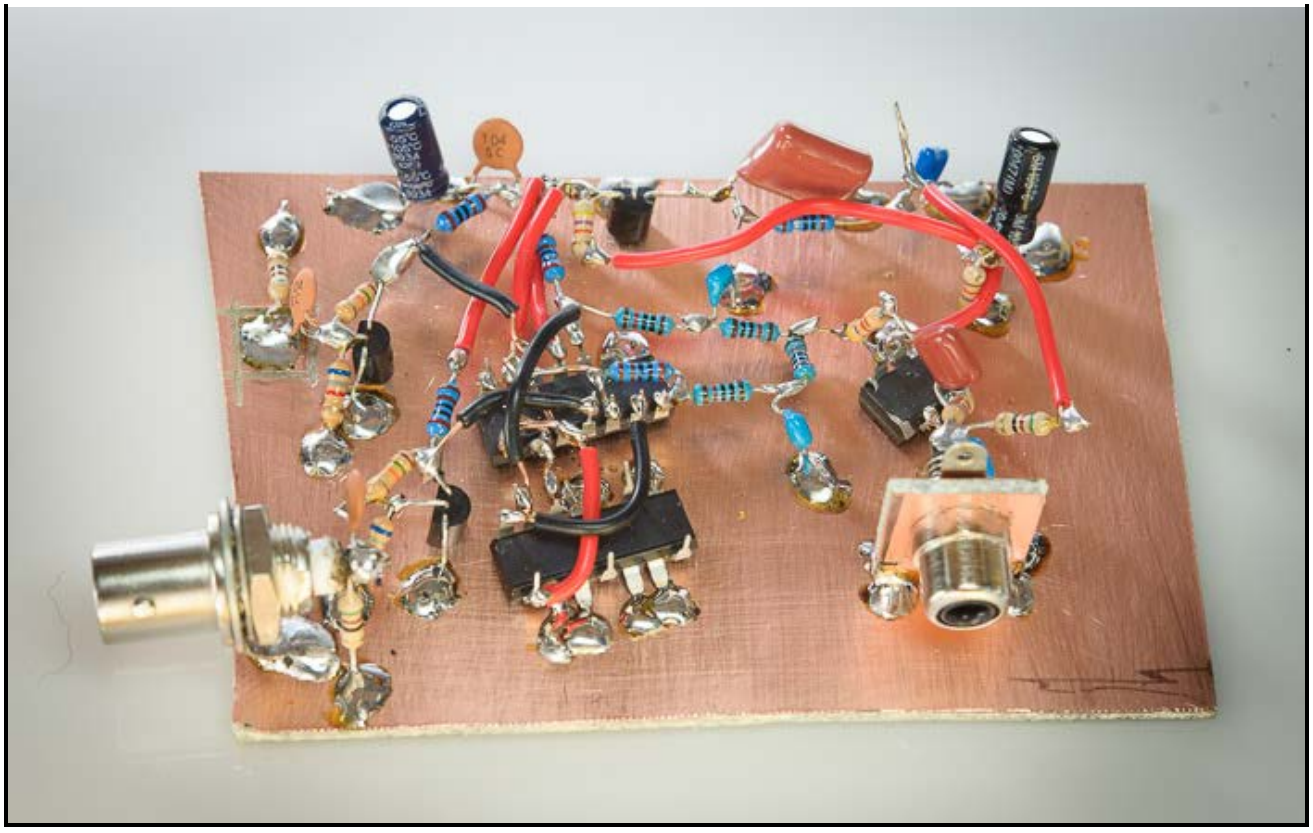
I inserted the 12 dB attenuator pad to keep my mixer RF port signal low to drop the mixer products amplitude down; further losses occur in the transformer. You can change this pad to whatever is required. I belong to the camp of builders who drive their transmit mixers with low-level RF signals to avoid messy outputs at the IF port.

Click for a breadboard photo of my initial bench tests with the mixer board. A 50 Ω resistive terminator shunts the main VCO output port during this testing. I temporarily insert BNC connectors along my development breadboards to measure output signals with my 50 Ω terminated 'scope, spectrum analyzer, or power meters and *rarely* measure RF circuits with a 10X probe.



Above — Phase-frequency detector and loop filter schematic. 51 Ω resistors terminate each 1.5 MHz input and drive two 2N3904 switches per EMRFD Figure 4.41. The loop filter design from EMRFD works as described, however, if you make a loop filter for a different circuit, casual copying goes out the window. Engineers design their loop filters according to factors including the crossover frequency, VCO gain, the N-division for the loop, etcetera *with software.* Some people and companies offer such software on the Web.

My loop filter 0.01 µF cap was a 1% polyester capacitor, although Wes specs a 10% tolerance in EMRFD. No cheapo ceramic bypass caps here please.



Above — Phase-frequency detector and loop filter breadboard. Click for a photo of the scattered, ugly, working boards on my workbench. Many prototypes look like this on our benches, however, sometimes, they work perfectly until we stick them in a box! Do you relate? Each oscillator belongs in its own metal box with strong bypass and decoupling networks (feedthrough caps reign supreme here) since the 3 oscillators might decide to party together and create havoc.

Above — My VCO frequency with the 1.5 MHz MTO set at full mesh. Since my MTO only tuned from 1.50 to 1.52 MHz, my VCO only tuned from

14.00 to 14.020 MHz, but that's easily fixed as I've stated. I'm very happy — it locked perfectly and my 14 MHz VCO stayed on frequency at the exact previously measured frequency drift of the 1.5 MHz MTO. When I connected the 14 MHz VCO to my counter without the PLL circuitry, it drifted willy-nilly.

Conclusion

The sense of awe and joy arising from locking a VCO on frequency won't be understood by many. The concepts and circuitry offers many possibilities. If the MTO and VCO exhibit low phase noise, short-term oscillator stability may be fantastic.

The 14 MHz VCO could be a 56 MHz VCO with sequential division by flip-flops to provide output at 28, 14 and 7 MHz with the 14 MHz portion going to the offset mixer. In EMRFD, Figure 4.44, Wes offers 14 and 7 MHz output by dividing the 14 MHz signal from the main power divider output port. The 7 MHz band is low-pass filtered to remove harmonic energy.

Wes extended this circuit by dividing the MTO by a hardware programmable frequency divider so that the difference from the mixer and low-pass filter is 170 kHz nominal. He uses this 'Almost Synthesizer' on the air for his QRP adventures.

While most builders will sensibly jump from an L-C VFO to a kit containing a programmed microcontroller plus a DDS or Si579, it's also fun to play with hardware to learn and ingrain synthesizer concepts + gain bench wisdom.

QRP — PosData for April 17, 2013

For a good read on the offset PLL, consider studying Wes' book **Introduction to Radio Frequency Design**, ARRL, 1994, page 320 and on. *This book is now out of print*. Wes ported the PLL active loop filter design program he wrote for IRFD from DOS to Windows in April 2013. Click and scroll for it. Thanks for this Wes!

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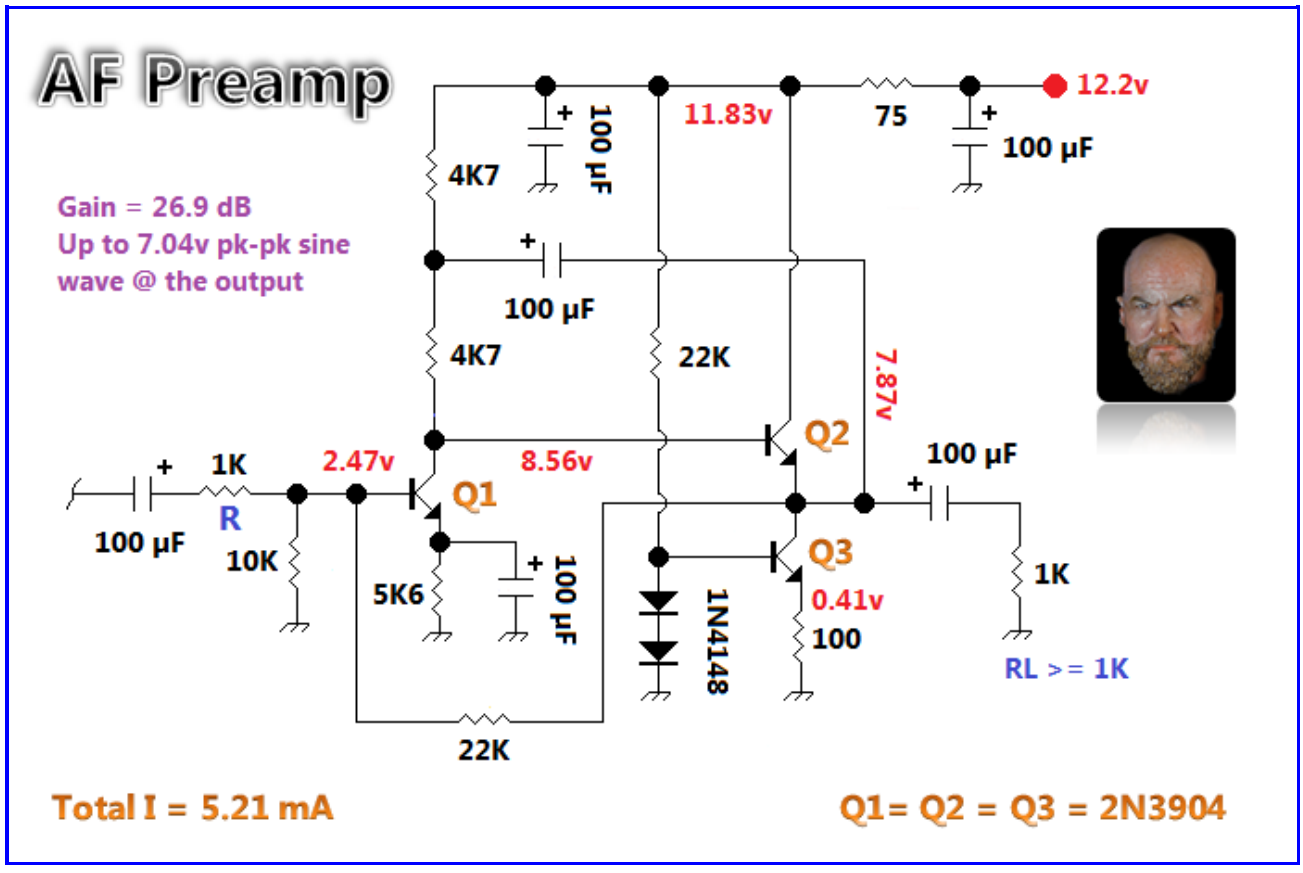
5. Boot-strapped Popcorn AF Feedback Pair



Above — I designed this AF stage for a builder from Indonesia; a popcorn AF shunt feedback amp based on the work of Douglass Self. Despite only drawing ~ 5 mA, this amp stayed clean until the output voltage exceeded 7.04 volts peak-peak on my test bench. Boot-strapping increases gain and lowers distortion in Q1.

Q2 buffers the Q1 voltage amp from external loading and increases gain.The Q3 current source boosts the load-handling capacity of the Q2 emitter follower. The input R can be raised to reduce sensitivity. The 1K output R could be a 5-10 K volume pot.

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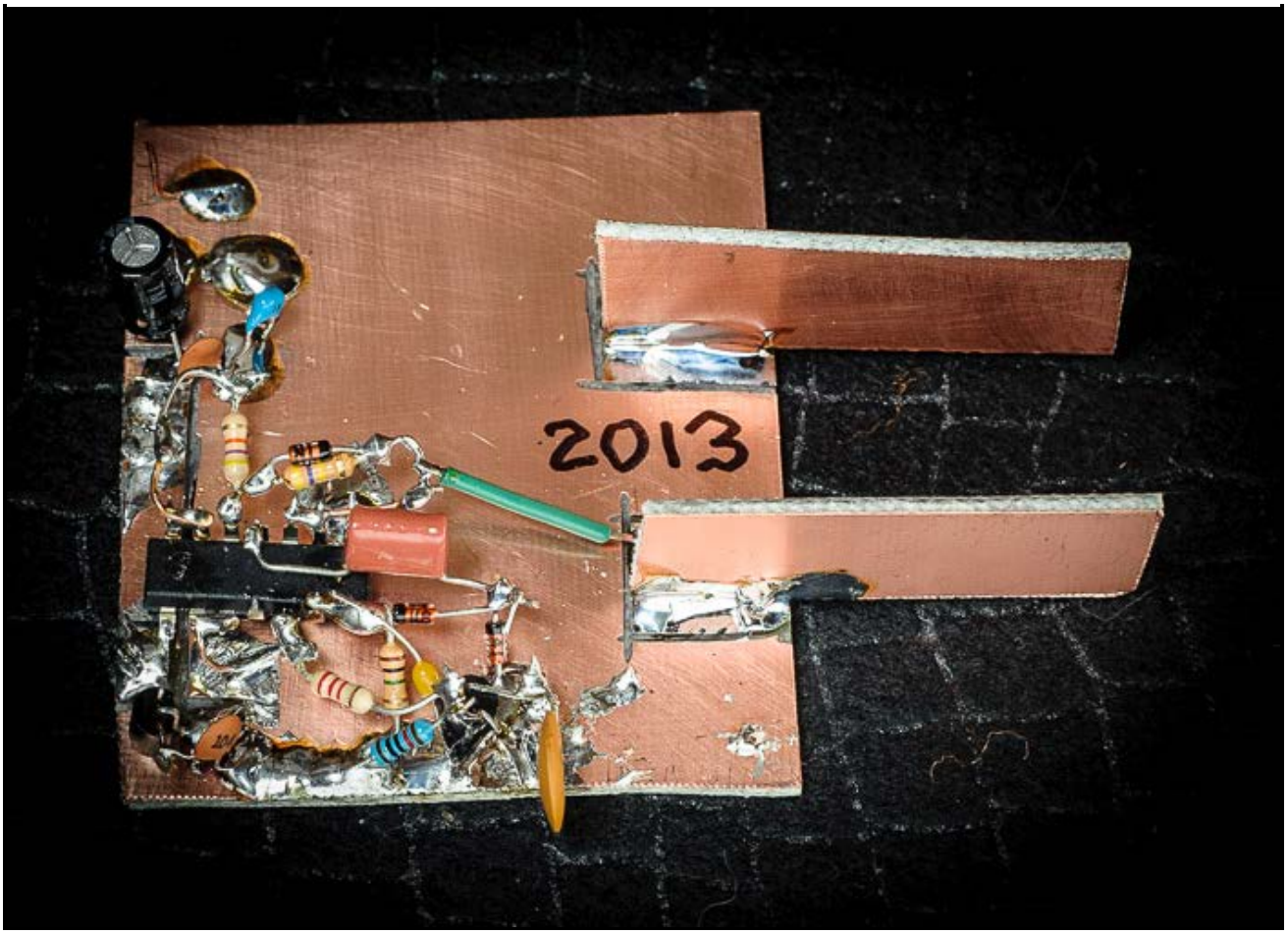
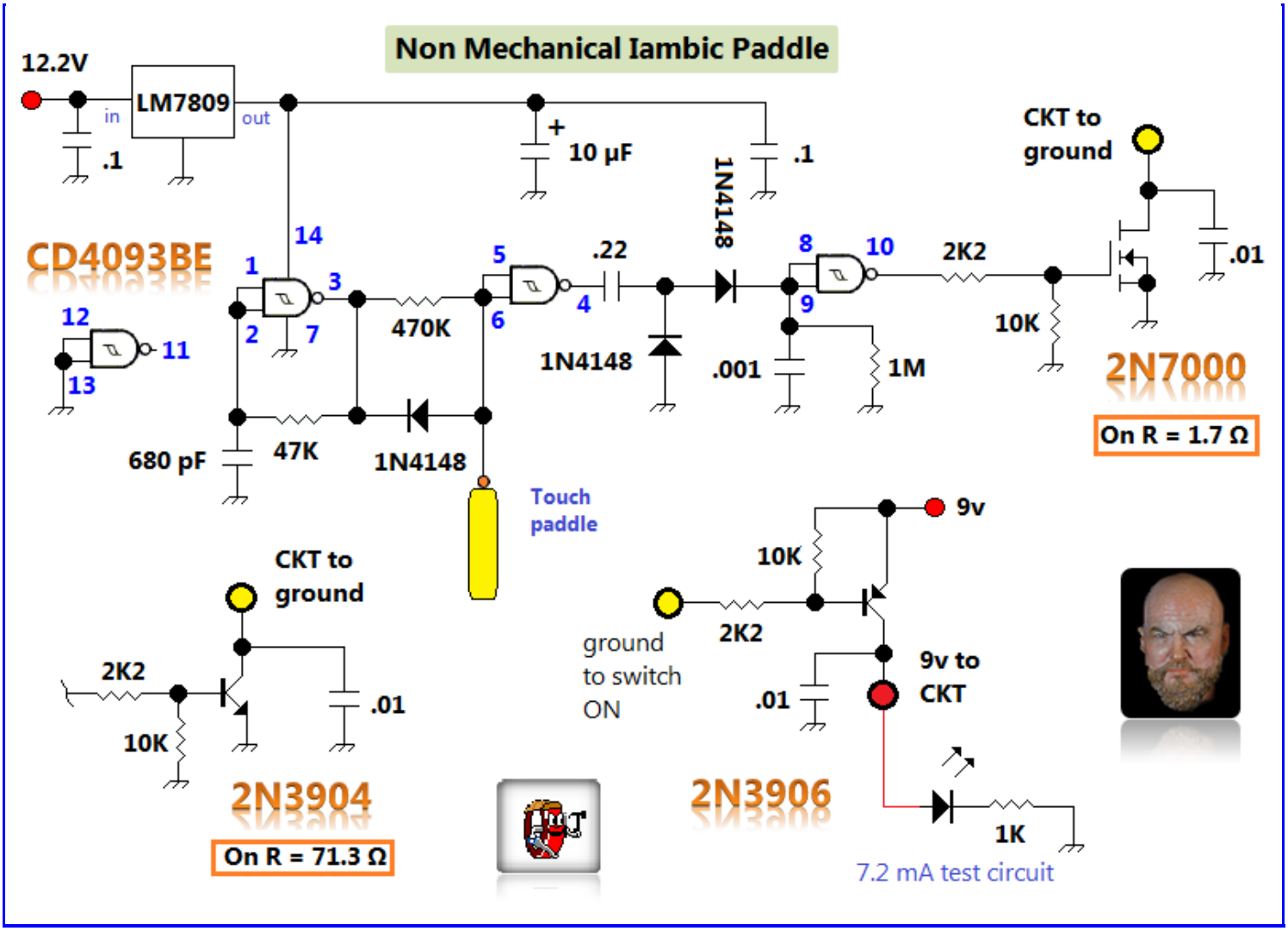
6. Non-Mechanical Iambic Paddle

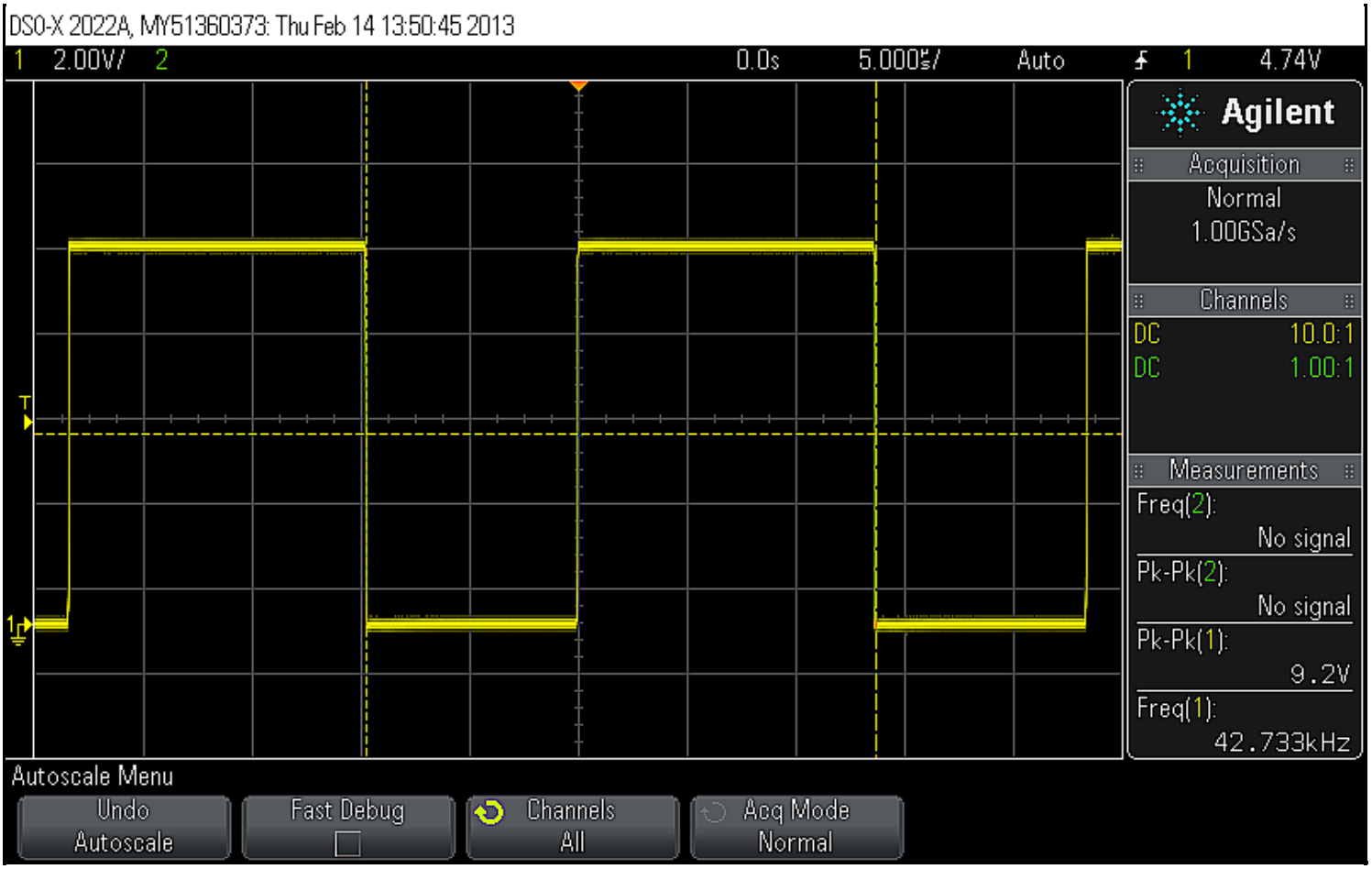


Above — The very ugly development proto-board of my half-done non-mechanical Iambic paddle. At some point I'll build the other half (the dah paddle switch) and press it into service. You might also use this circuit as a non-mechanical straight key.

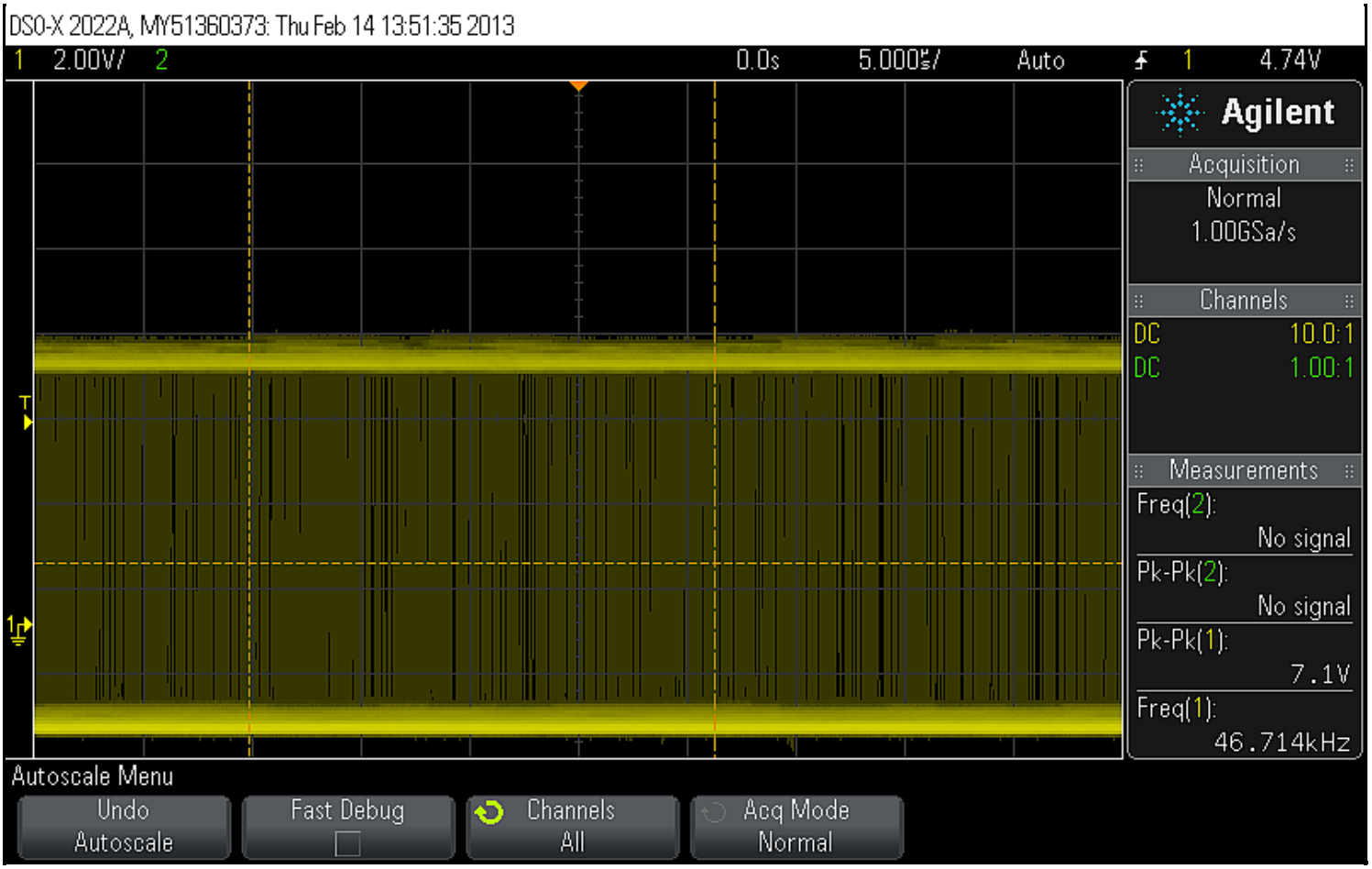


Above — The schematic for 1 of the paddle circuits. I compared the ON resistance of the BJT switch with the enhancement mode FET and the 2N7000 won: only the FET could key the continuity tester on my DVM.

You may extend this circuit with a 2N3906 switch for paddle-switched 9 volts (or whatever VCC you want). In the bottom right, I connected the PNP collector to an LED and flashed it for fun. The 0.01 µF capacitor on the switch drain or collector bypasses any RF to ground. With higher power RF, you may have to place a similar bypass cap in parallel with the shunt 10K resistor on the 2N3904/2N7000 base and gate respectively.



Above — A 'scope shot of the ~ 43 KHz oscillator generated in the first 4093 Schmitt trigger NAND gate.



Above — Here's the disturbed oscillator output *just* after the paddle is touched: this stops the signal at pin 5 and 6 of gate 2 and kills the AC output at pin 4. The DC voltage across the 0.001 µF cap discharges through the 1 megohm resistor pushing pin 10 HIGH to turn on the 2N7000 (or the 2N3904). Normally, pin 10 is LOW since the rectified output of the undisturbed ~43 KHz oscillator goes to both pin 8 and 9.

A fun circuit for a Saturday afternoon...

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7. A Journey Above HF

This project began as a 14 MHz low-noise amplifier build, but ended up with me learning more about SMT breadboard techniques and suppressing spurs. A short exploratory/descriptive account of my bench journey plus some photos follow.

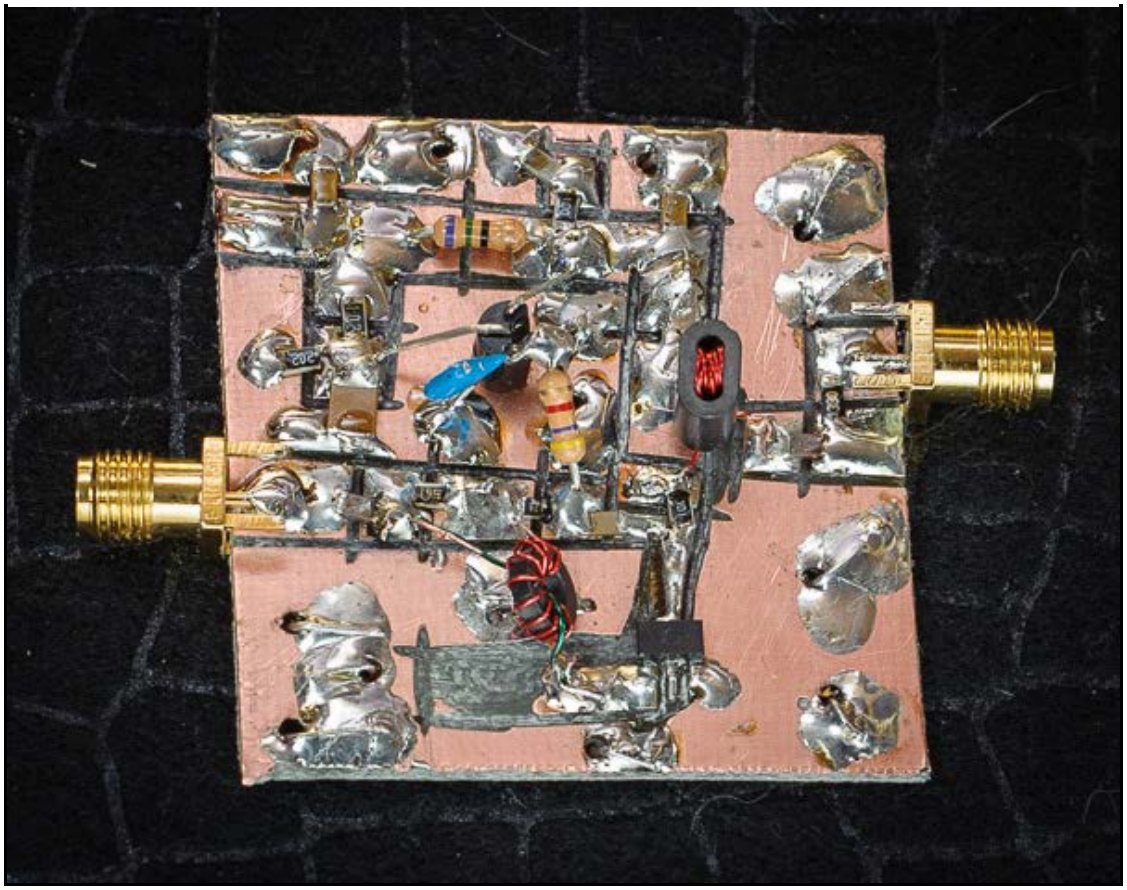
Above — I'm slowly adding SMA connectors and pieces. Since modern consumer digital network engineers use them, they're abundant and often rated from DC to 18 GHz; more bandwidth than I'll ever need.

I'm also building with evermore SMT components and just love it. Through-hole (I prefer to say hole-through) stuff continues to disappear like lemonade on a hot August afternoon.

Above — The schematic of my version of Victor 4Z4ME's feedback amp (FBA) as tested at 14 MHz: he emailed me a paper and provided some online support. Click for another version from December 2012. Typical noiseless FBAs suffer from poor reverse isolation, however, Victor runs the collector to base feedback through an asymmetrical 3 dB power combiner/splitter that boosts port isolation, defines the gain, plus sets the input and output impedance.

Strong virtues of the asymmetrical power splitter — fully utilized in this design include a very low loss on the input side (the 1 turn side) and a much higher loss ont the feedback side that allows the feedback to defines the input impedance on one side while exerting a negligible impact on noise figure and dynamic range on the other side. Victor measured a noise figure of 1.5 dB using a MRF586 BJT. Thanks to Victor for the information and design.

For strong IMD properties, I ran 50.6 mA total stage current into a gorgeous, low-noise, NE46134 NPN transistor with a fT of 5.5 GHz. Using VHF-UHF techniques, I built with mostly SMT parts on 2–sided board using copper wire vias to connect the 2 copper surfaces. I discussed the wrap-around bias technique in 2011 as number 1.



Above — My prototype breadboard with dremel cut islands for soldering the size 1206 or 0805 SMT parts, plus a few hole-through items. Carving an island for the SOT89 transistor package proved difficult, but even I (a challenged dremelist) did it.

Woe to Oscillations: Like misplaced car keys, oscillations may remain hidden unless you search for them. Often, the only difference between a proper oscillator and a regular amplifier is we *want* the former to oscillate. To check for instability, we might use our high bandwidth scope, or a spectrum analyzer, but many will have to find spurious RF with basic, DC - HF bandwidth test equipment. In any case, just do your best. To some extent, unwanted oscillations are the *elephant in the room* that few talk about. Well, it's okay to think, talk and feel some emotions about them.

Sure enough, when I connected a 14 MHz signal to the amp's input and a 50 Ω terminated 'scope to the output to measure gain, my sine wave had 2 or 3 others on top of it. In the 4Z4ME amp, the PNP bias transistor can be a source of AF to HF oscillations.

Victor wrote: "The circuit has a low frequency amplifying loop that goes through both transistors. The PNP transistor does not invert the signal (it is a common base amplifier) and the RF transistor inverts so it is a loop with 180 degrees phase shift (negative feedback). The various decoupling and RF coupling capacitors in this loop add phase shift on this low frequency loop. If the accumulated phase shift adds to an additional 180 degrees and gain is larger than 1 you have oscillations. The simplest way to solve it is to make one of the capacitors very large so it will add only 90 degrees phase shift but it will drop the gain at the higher frequencies where the other capacitors start to add phase shift to be less than 1 so there are no conditions for low frequency oscillations. This technique is called "Dominant Pole". That's the reason that I suggested to connect a very large capacitor to the PNP transistor".

I found my oscillations disappeared with a 0.1 µF collector bypass cap on the PNP (Cx on my schematic). The 0.1 µF cap on the PNP collector was critical – a 0.22 µF failed to work, as did a .001 µF --- but a 0.1 µF held it stable. In another 2N2222a-based 4Z4ME amp with 0.01 µF input and output caps, it took a 10 µF capacitor on the PNP collector to snuff out some ~766 Hz oscillations.

We don't use a wrap-around PNP bias with our RF oscillators — that's asking for trouble.

I aso measured oscillations at ~ 372 MHz with my spectrum analyzer. A collector 10 Ω R killed these UHF oscillations and after that I saw no spurs from .001 to 1 GHz. (I should have made the dremel cut right close to the NPN collector for the 10 Ω resistor. I hoped there were no oscillations above 1 GHz because I can't measure them.

Finding oscillations:

Many builders lack a spectrum analyzer, let alone 1 that goes up into UHF bandwidth. I'll share a few tips I've learned on the bench that don't require expensive test gear:

A 10X scope probe on the drain or collector of an amp may sometimes reveal oscillations up to the maximum 'scope bandwidth — set your 'scope vertical scale for high sensitivity. This provides direct measurment of oscillations.

Indirect methods to infer unwanted oscillations also lie in our armatorium. I learned this trick from Wes: Place the circuit under test in your normal gain measurement set up with an oscilloscope. Then vary the DC power supply voltage slowly and smoothly — your measured 'scope voltage changes should also track slowly and smoothly. You may see an AC voltage jump as the amplifier goes into and out of oscillation with the DC power supply tweaking. After finding this oscillation caused AC voltage spike, you work to remedy it with a variety of means such as better bypassing, changing bias voltages, shielding and locating breadboard errors.

Sometimes if you put your finger near the active part while watching the bias voltage or current you may see the bias jump around *if* oscillations are present.

My final indirect oscillation busting technique: If you measure the specified/expected gain and return loss on the input/output port, this *may* signal your device is stable — I've noticed this with MMICs where I saw oscillations on my SA, stabilized them and *only then*, measured the expected S21, S11 and S22.

Sometimes, eliminating a hot part proves the best fix! In 2012, a new builder wrote to say that he soldered in a Mini-Circuits DC - 6GHz MMIC; the ERA-2SM in SOT-86 as a buffer for his 3.5 MHz VFO. Anyhow, in the photo were long leads plus no decoupling resistors etc. It sounds like the circuit behaved hyperreactively and *vibrat*ed in spasm. The cure was to eliminate the microwave part and put in a hycas amp built with a J310 + a 2N3904 — we encounter risk when plying the latest, hottest, super-high fT amplifiers sold on eBay with casual abandon.

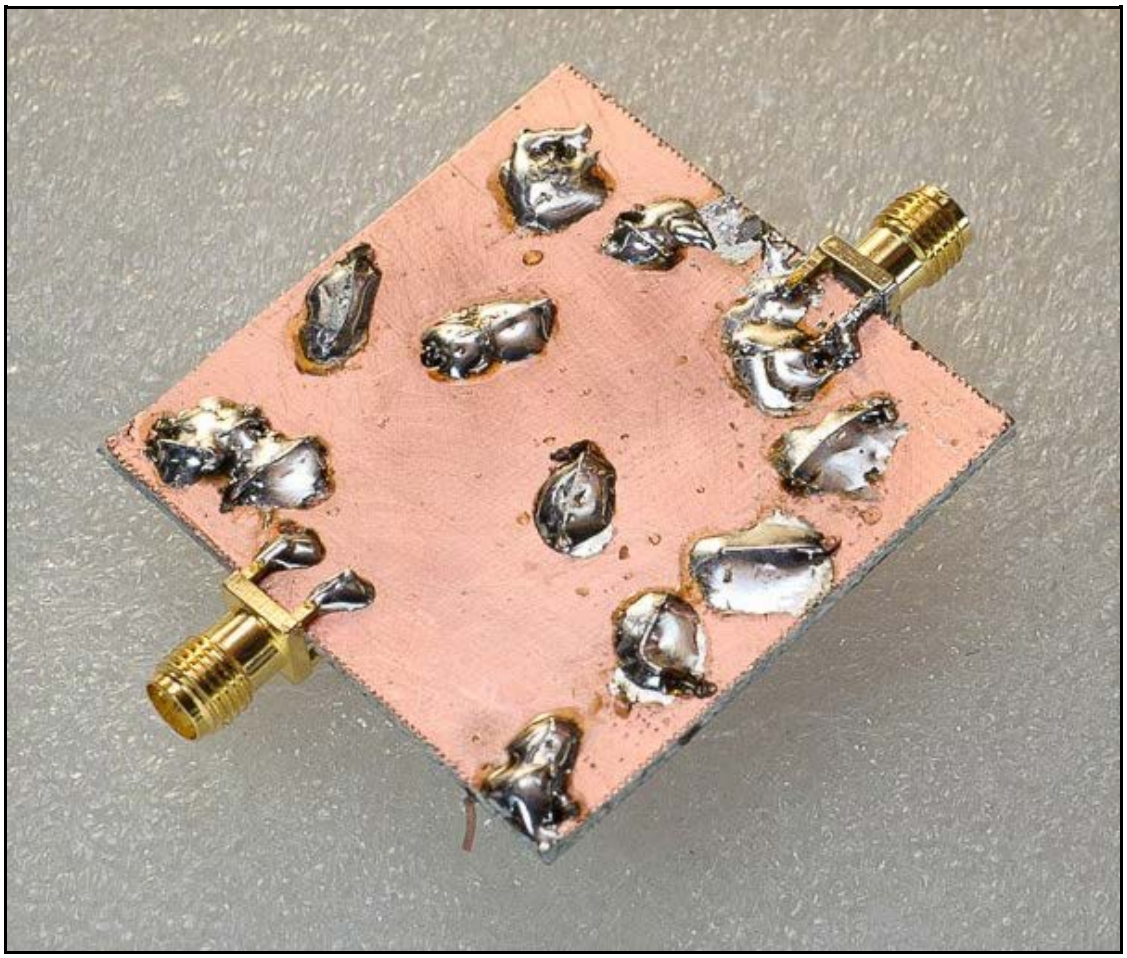
Practice makes perfect. if you believe learning is experiential and build to learn, you'll learn to build.

Finally, as an amateur, I struggled to choose a SMT ferrite bead and after reviewing many datasheets and application notes I ordered a size 0805, 800 mA part with 120 ohms Z at 100 MHz and its peak impedance at 340 MHz. I'll let you know how that works out.

RF Bypass on our DC lines: As possible, we ought to provide a broadband RF bypass to provide a low impedance to RF from low frequency up to the maximum frequency wherever our FET, BJT, MMIC, etc. operates. For example, you can't just swap a higher gain BF998 (1 GHz) for a 40673 (VHF) dual-gate MOSFET and expect the same stability and bypass requirements can you? At the very least, I bypass G2 of the BF998 with a size 0805 0.01 or .001 µF SMT capacitor and the drain with RF bypass good for 1 GHz.

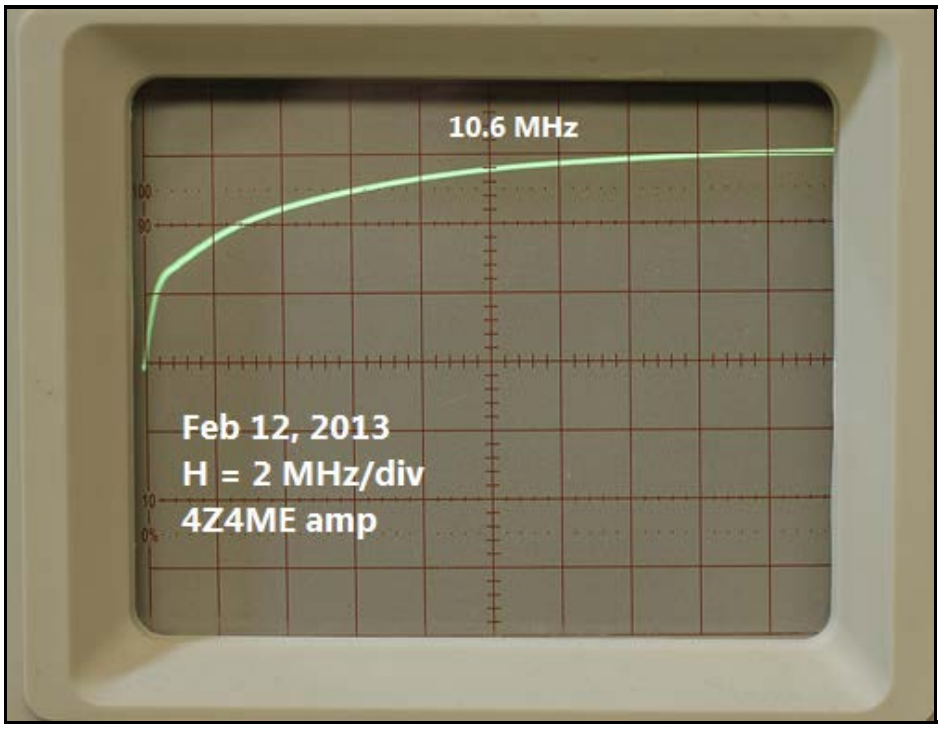
Wideband RF bypass may solve oscillation issues too.

I tried to apply a broadband bypass in my breadboard, although it gets extremely difficult to think about bypassing RF at > 1 GHz for the QRP homebuilder. Our hobby should include reflection and proper intention at the very least.

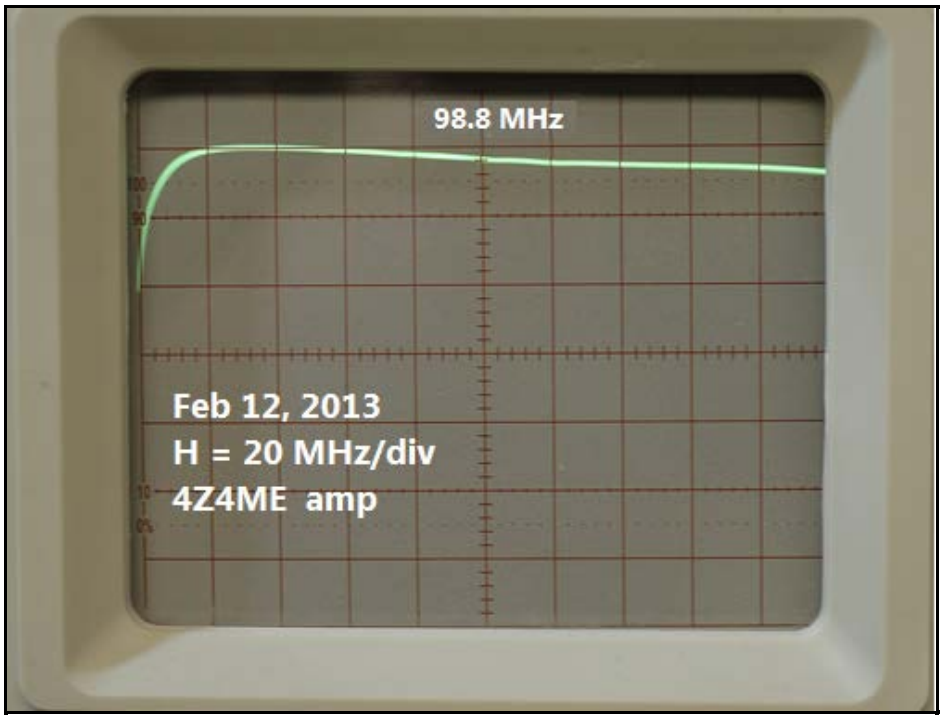


Above — A photo of the bottom of my breadboard showing the vias. I made mistakes: we should try to keep the via holes as close as possible

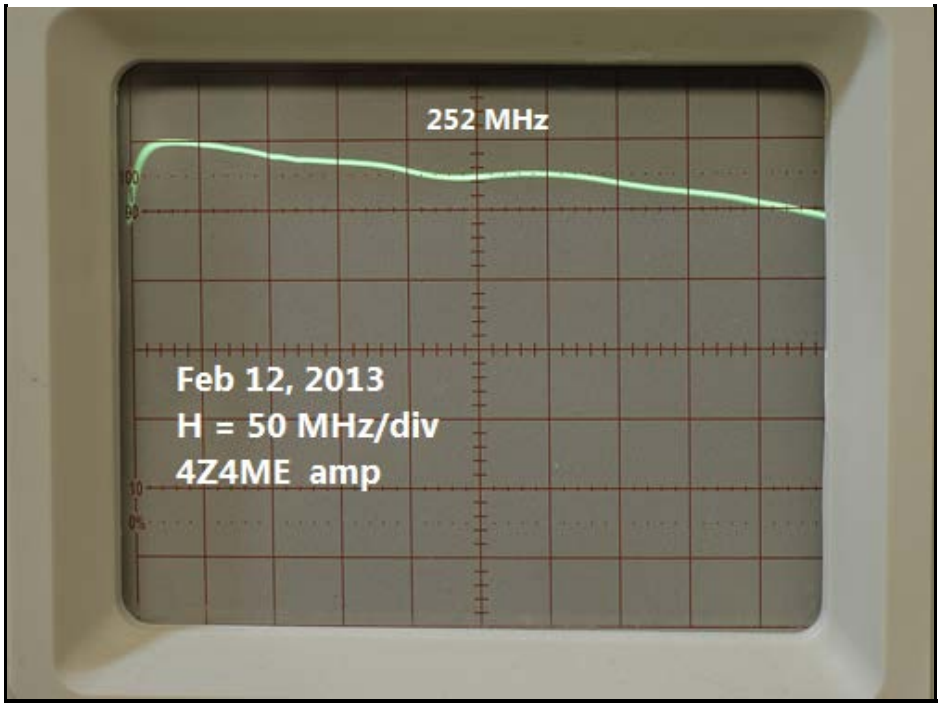
to all bypass caps, my 10 Ω collector snubber resistor, collector port, or whatever we need to put at RF ground. The vias connect circuit areas to the large area, low impedance ground plane to minimize inductance. We should also try to place bypass capacitors as physically close to the pins of whatever we're bypassing.



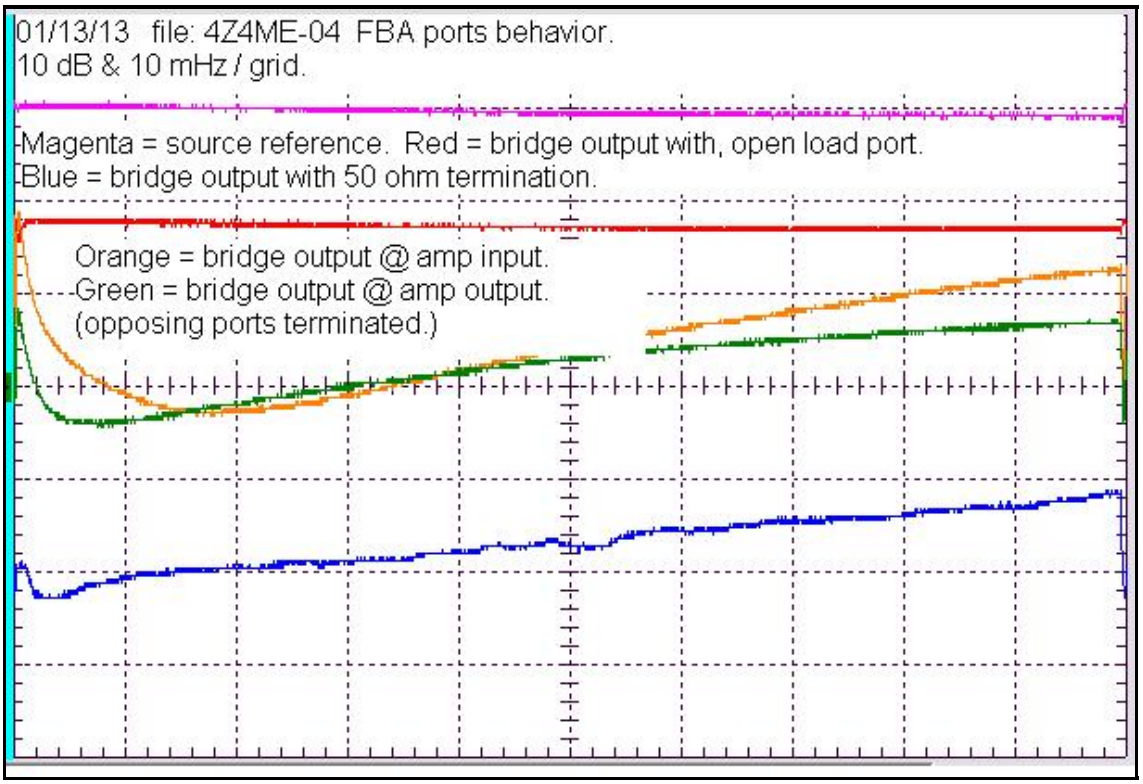
Above — A TG + SA sweep of the gain of my 4Z4ME NE46134 FBA from ~1 to 20.6 MHz. Each vertical quare denotes 10 dB. Each horizontal square = the value specified in the photo. In this case; 2 MHz per division. Maximum gain was ~16 dB.



Above — A sweep from ~ 1 to 200 MHz. This would also make a good 6 meter band amplifer or ????.



Above — A sweep from ~ 1 to 500 MHz.



Above — Bob K3NHI made and swept a 2N5109 version of the 4Z4ME amp biased for ~47 mA emitter current. Here he swept return loss at the amp's input and output from about 1 to 100 MHz. The output return loss of my build was down, however, I didn't own many SMT resistors between 33 and 68 Ω. For example, if the output impedance at the collector is 10 Ω, then the series resistor should be 40 (39) Ω.

I've found that in my FBAs, changing the current and also the transistor type (2N5109 , 2N2222a etc.) also affects the input and output return loss. At HF, it's possible to measure RL with a simple bridge, so optimization is possible.

I learned a lot by building just 1 amplifer and discussing my findings by email with friends. Hopefully the next version I make will show improved understanding and skill.

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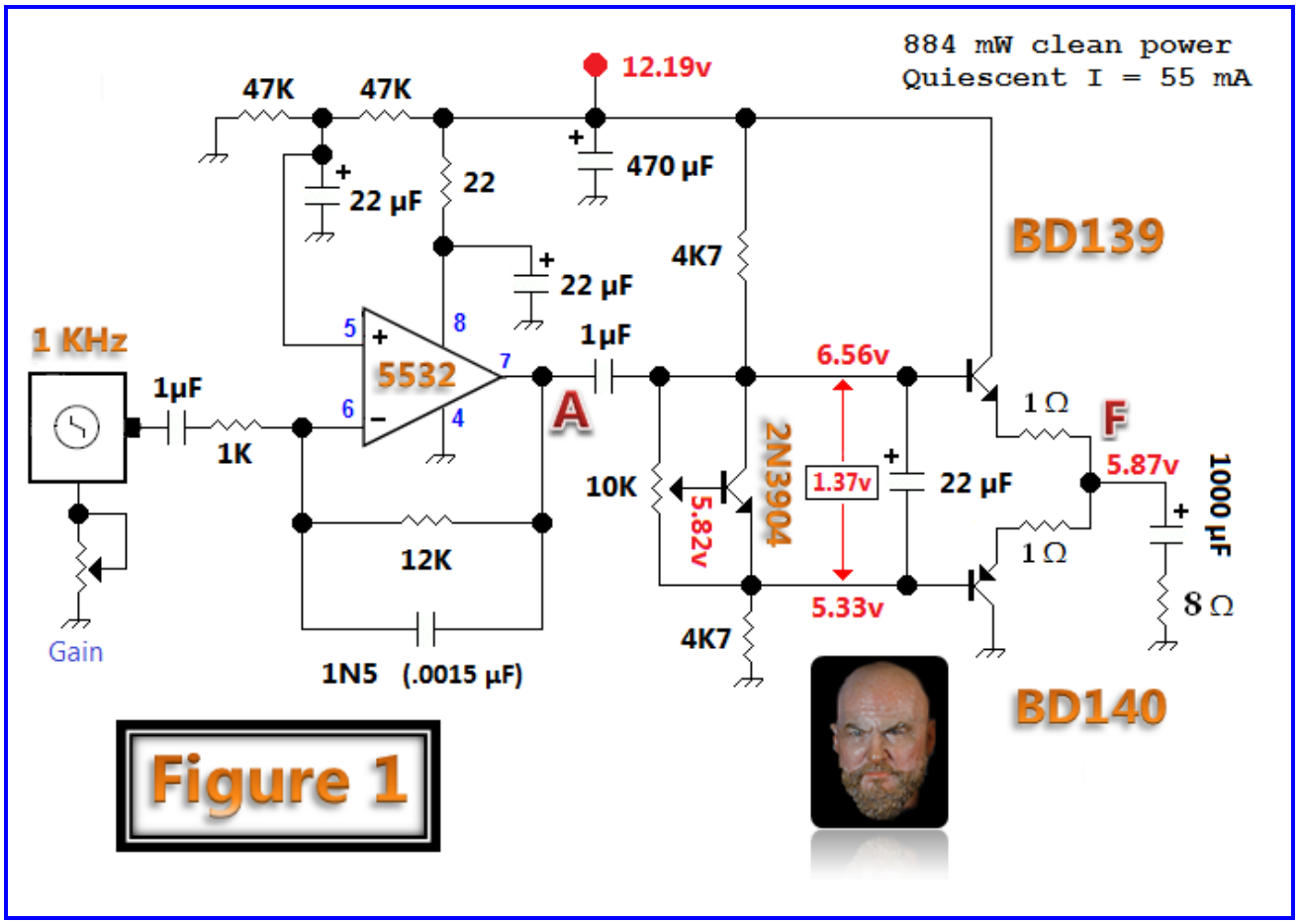
8. Popcorn AF Amplifier for Receivers — Reprise

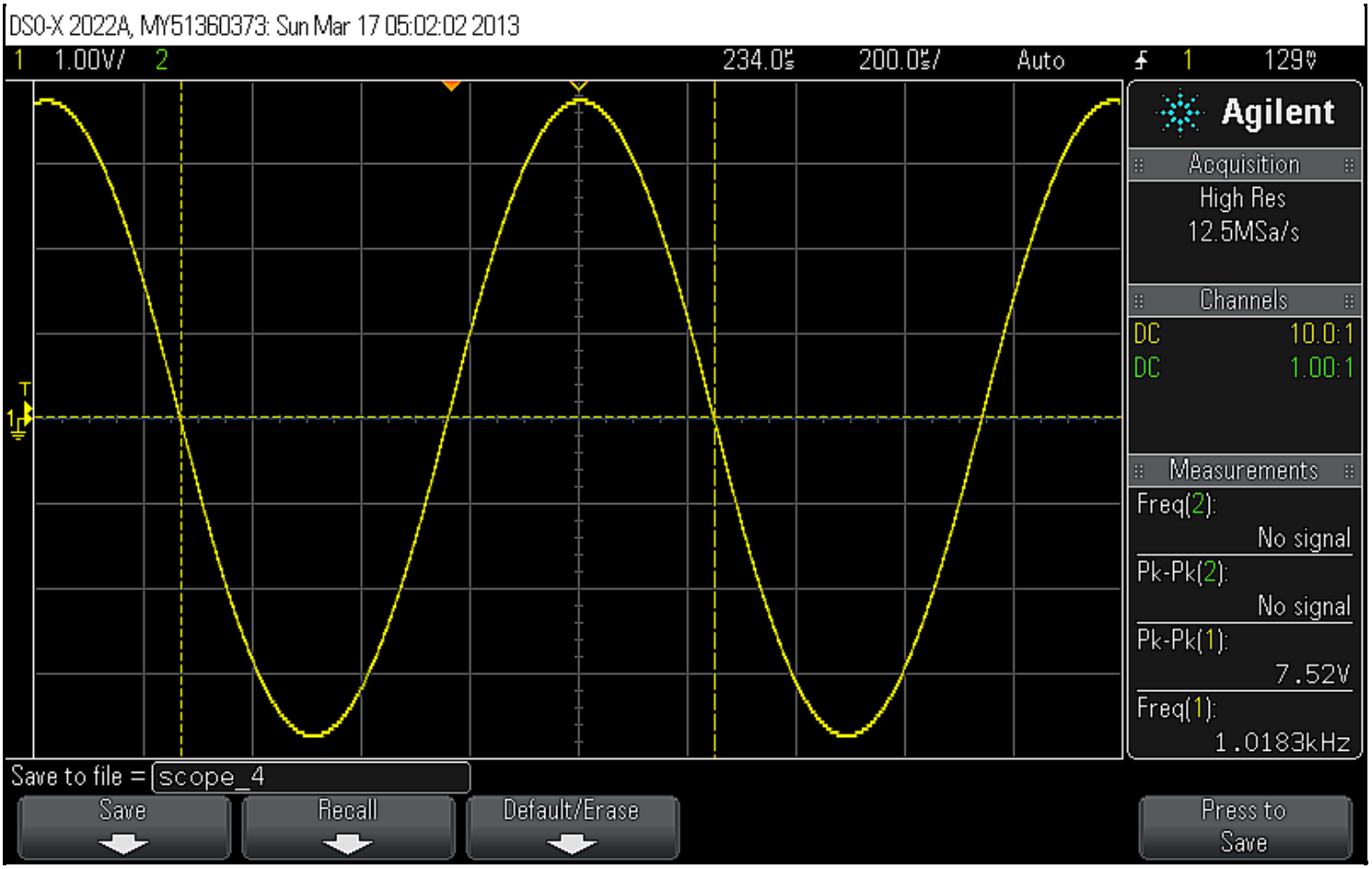
I've worked on a popcorn audio power amplifier (PA) since 2008 and offer my latest experiments. There's only so much you can do with a single-supply 12 volt AF power amp, but I enjoy improving my circuit.

My power measurement technique is shown as Figure 4 here. To enhance versatility, the following PA's may be coupled to whatever preamplifer you choose. In all cases, I drove the power amp stage with a 5532 op-amp voltage amplifer. The power followers were biased with a 2N3904 amplified diode (also called NPN shifter bias amplifier, or DC level shifter) rather than just a pair of series diodes, since this allows you to dial in just the right amount of bias as you watch the AC signal in your 'scope. I wrote a tutorial that explains how to bias complimentary-symmetry power followers in 2008: Click for the link.

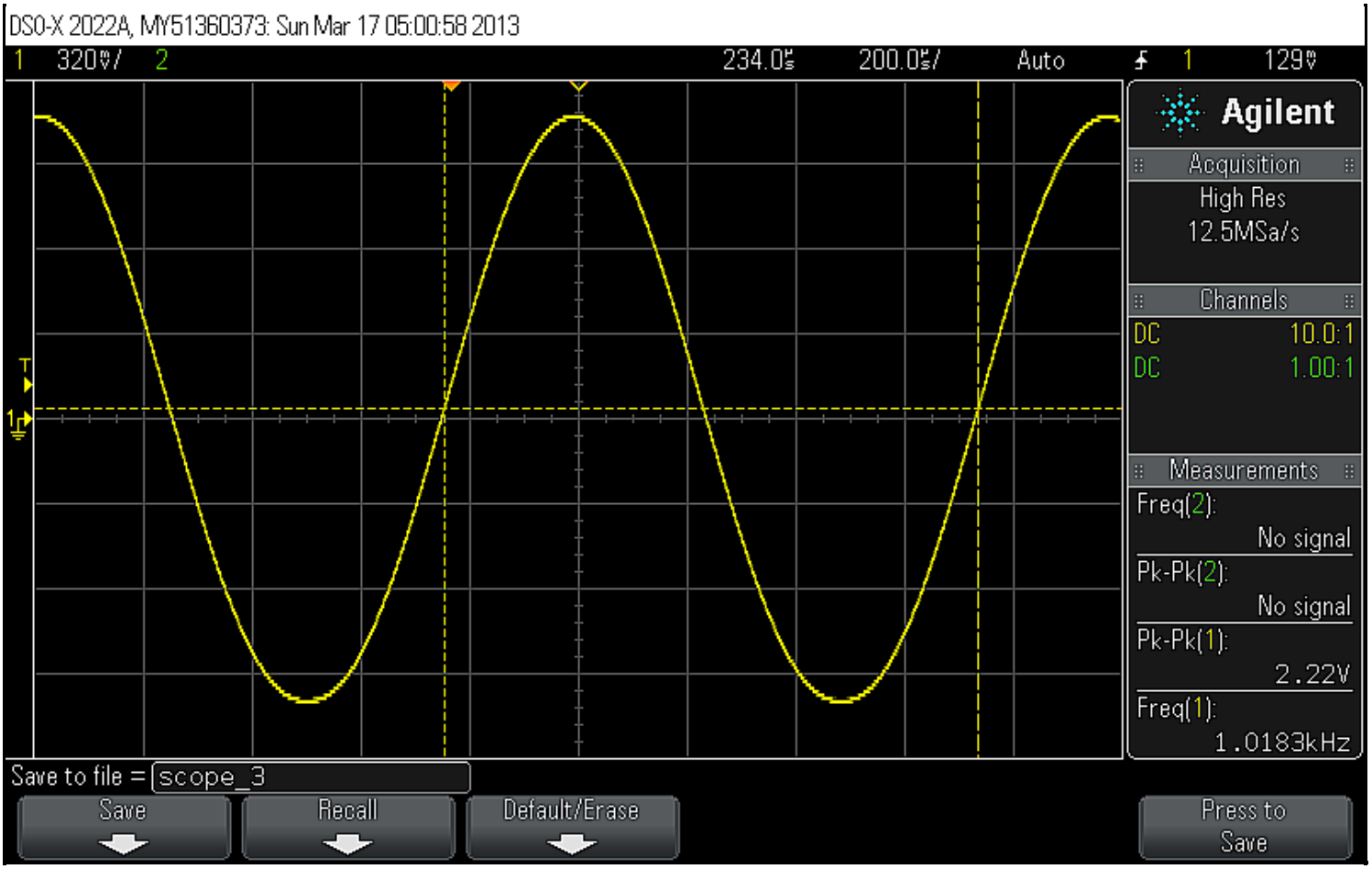


Above — Figure 1: A popcorn AF power amplifer in full bench *test* mode. Measure the AC with a 10X 'scope probe across the 8 Ω resistor and the DC voltage and current with a multimeter. A distortion analyzer proves useful, but not essential for popcorn circuitry. I also listened to each amplifer connected to a line-level cassette player and an 8 Ω, 15 cm speaker. A 4 Ω speaker doubles the maximal clean power, but I don't own any and stuck to 8 Ω.

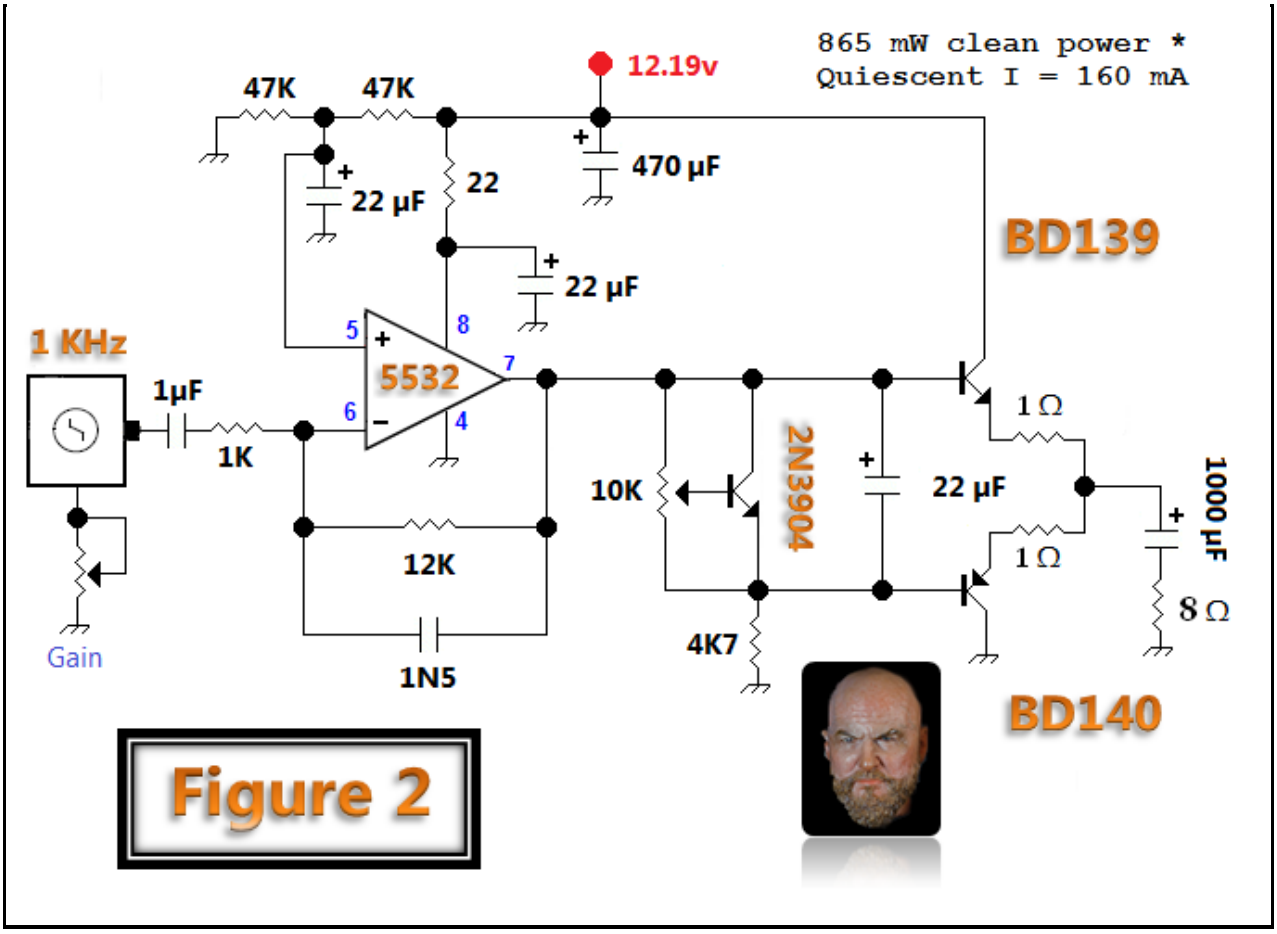
Containing no negative feedback, the power amplifer stage runs from the red-colored designator points **A** to **F**. You can AC or DC couple point F to your preamplifer stage as required to apply negative feedback.
As mentioned, you can use the 5532 preamp shown with any reasonable gain (i.e. change the 12K resistor), or opt to replace it with your own design. A low output impedance amplifer best drives the power stage.



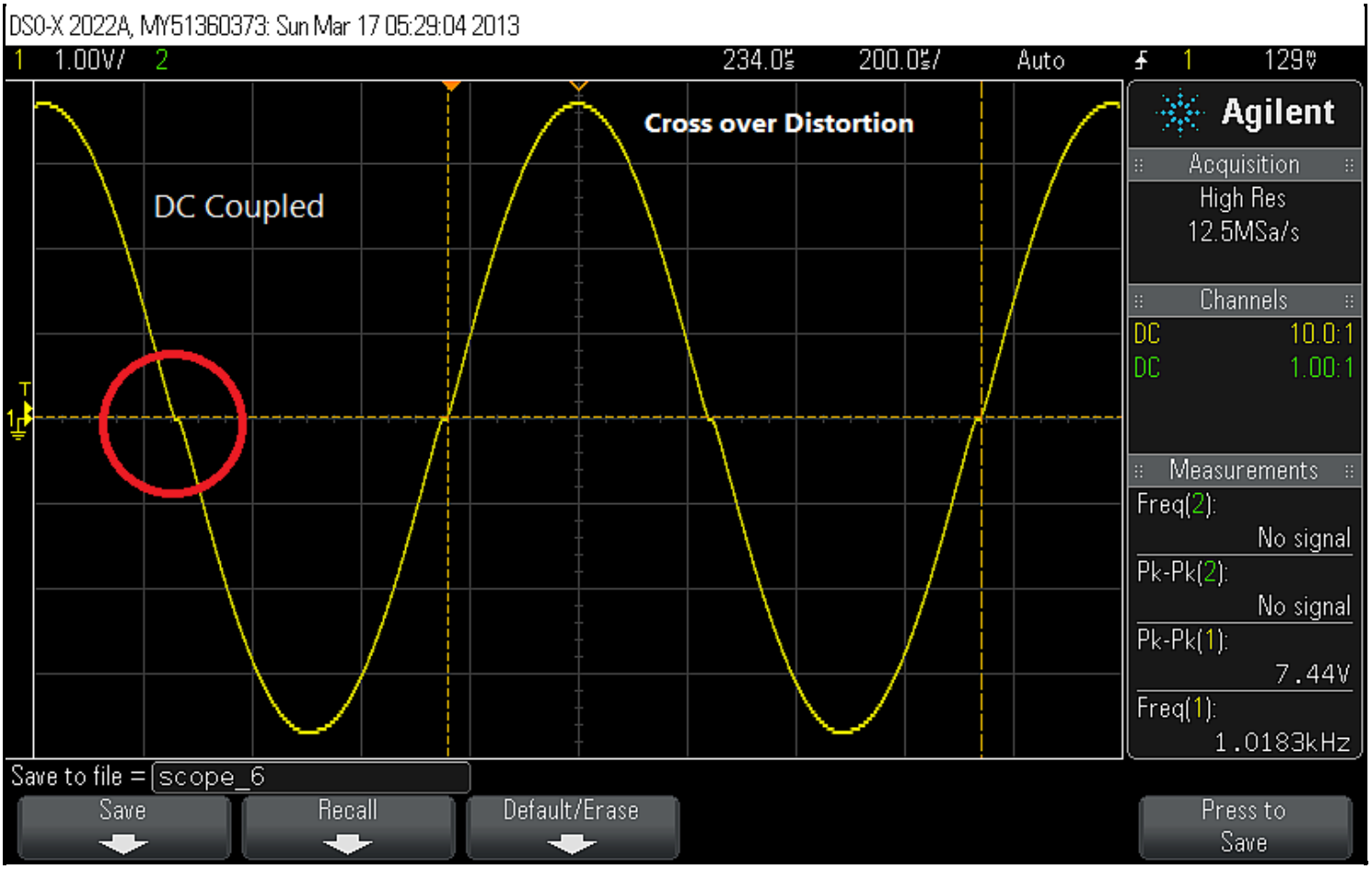
Above — The output of Figure 1 in my 'scope driven to the maximal pk-pk voltage just before distortion begins to appear. Obviously this task is somewhat subjective, however, allows comparison of the amps you build on your bench.



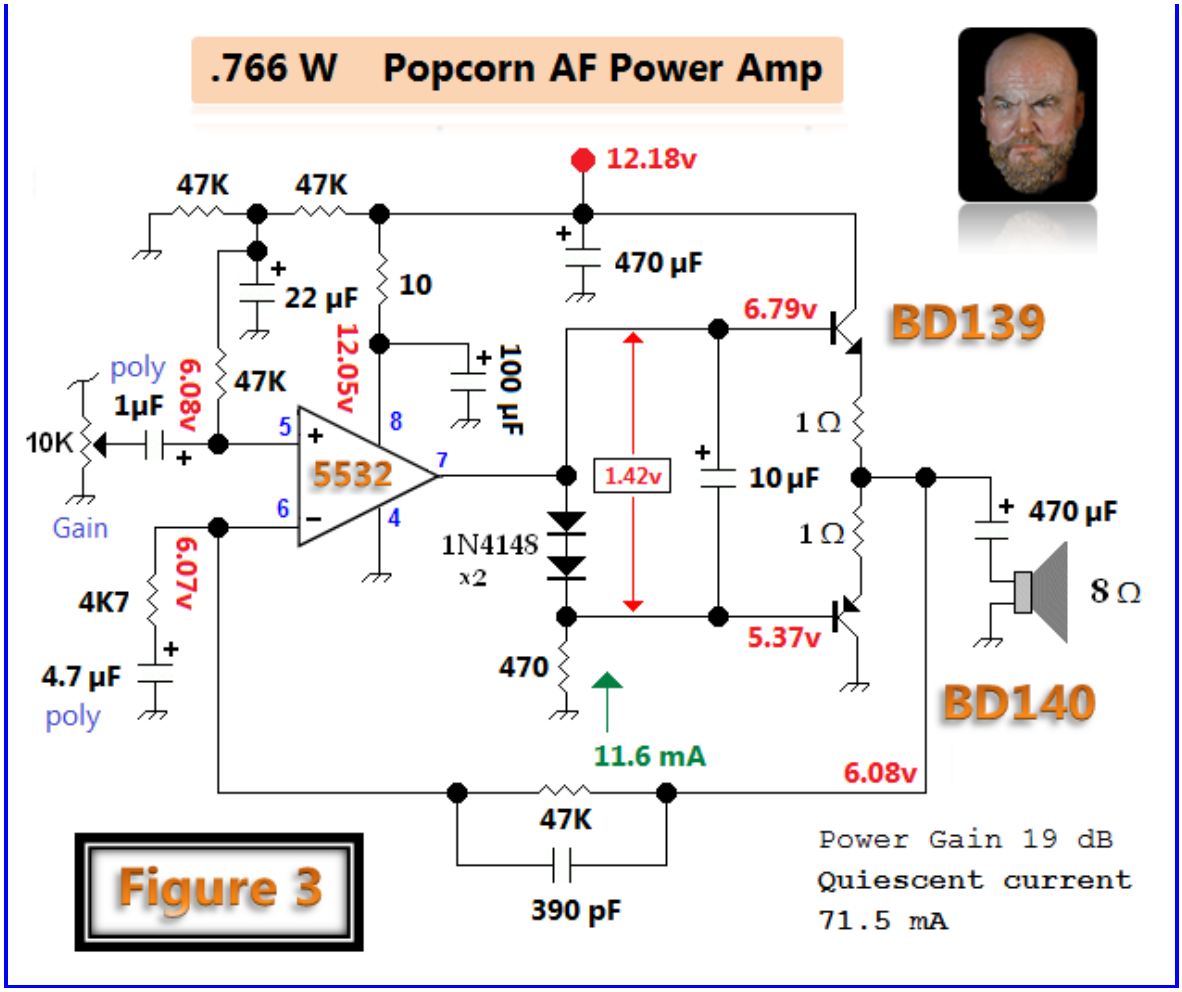
Above — A 'scope screen capture with the 22 µF level-shifter filter capacitor from Figure 1 removed. Look what happened; the maximum clean signal fell from 7.52v pk-pk to 2.22v pk-pk. That capacitor is essential to get the maximal possible headroom.



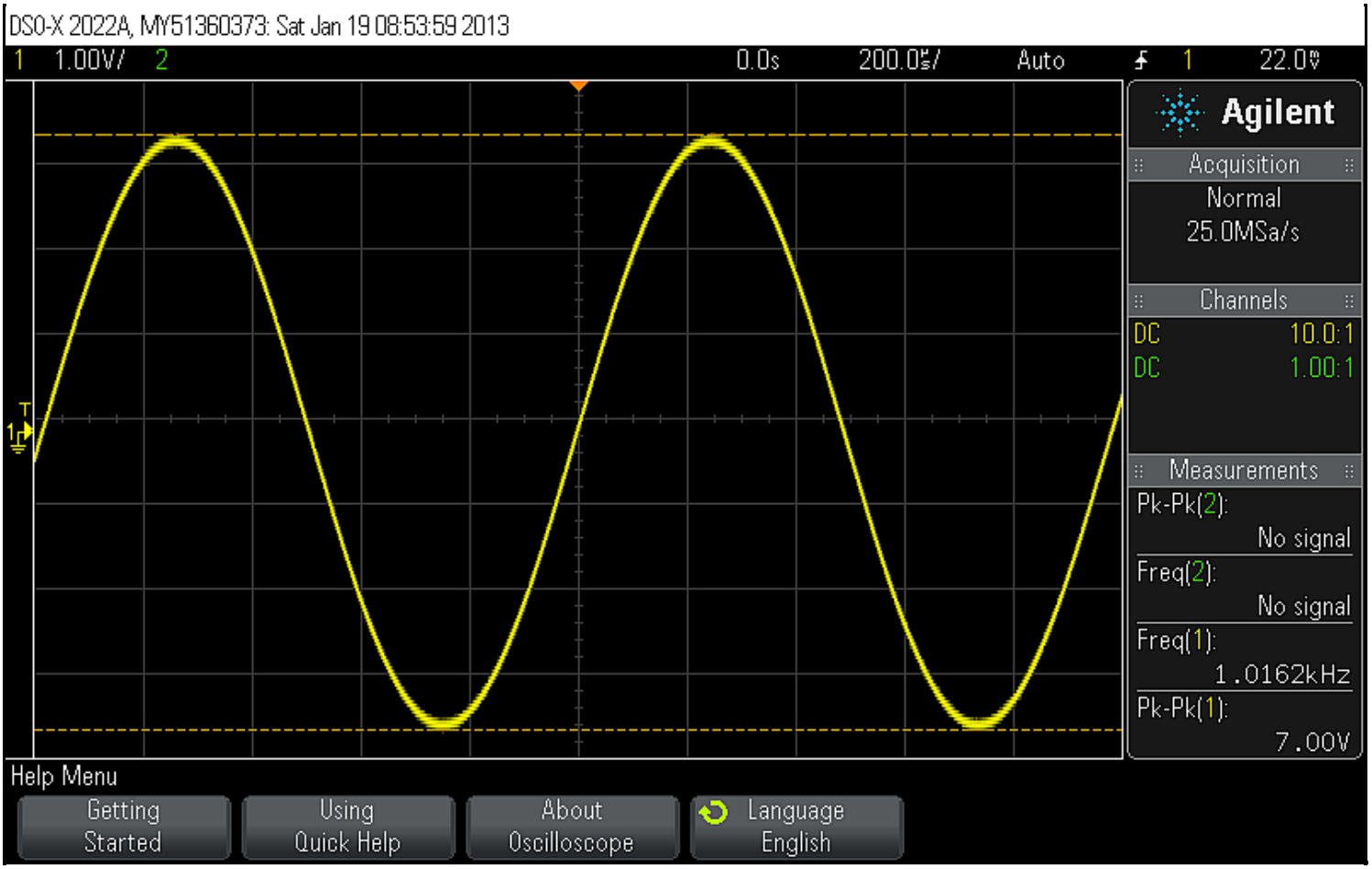
Above — Figure 2 is Figure 1 with the op-amp DC coupled to the level-shifter. I tested the circuit with and without the 4K7 resistor connecting the base of the 2N3904 to the DC supply: it didn't boost the amplifer headroom, nor reduced crossover distortion, so I removed that R.



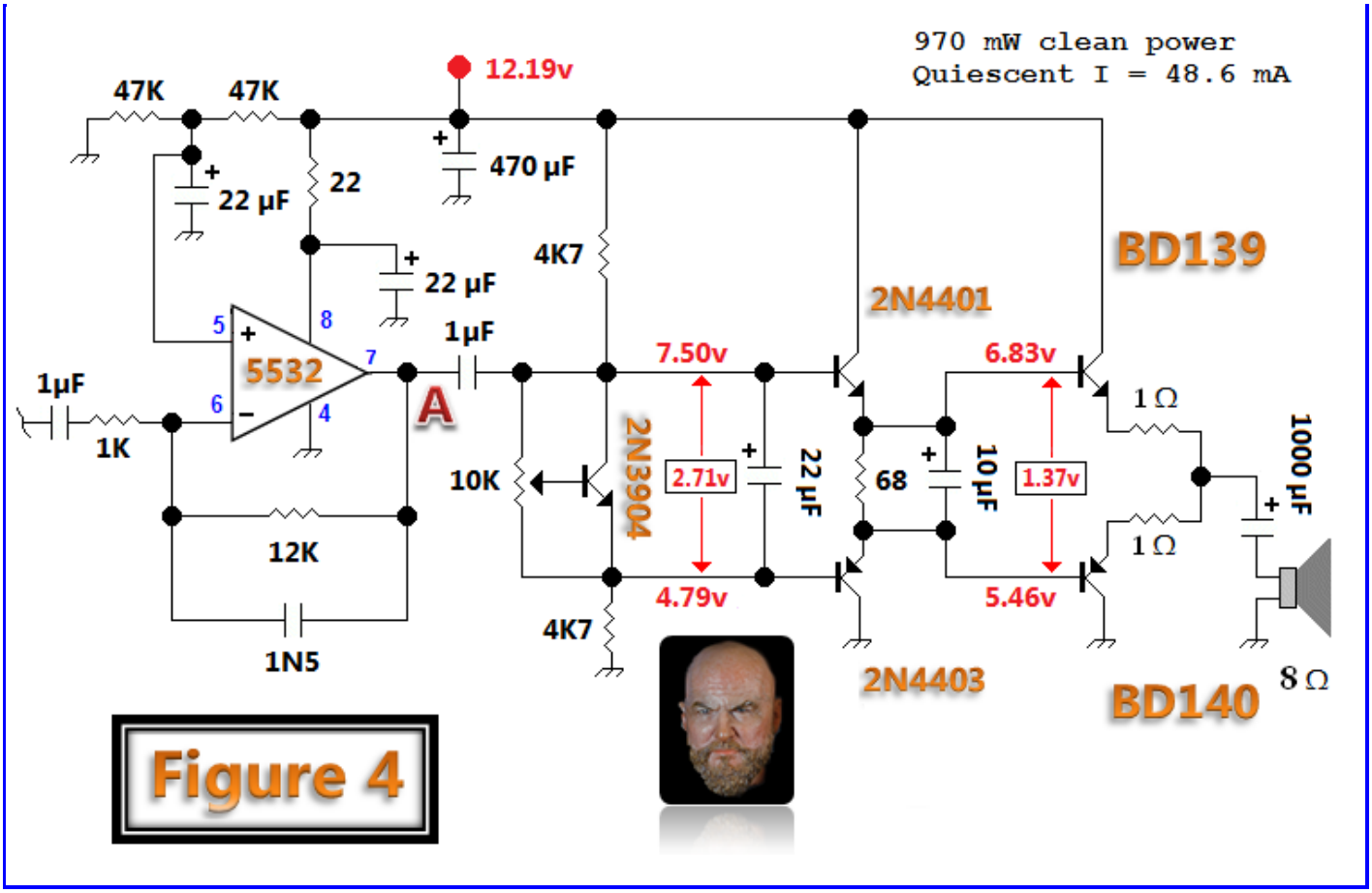
Above — The Figure 2 amplifer 'scope tracing. At *maximum power*, crossover distortion appeared and I've seen this before. Likely, there is not enough base drive to keep the power followers forward biased. By adjusting the level shifter, I almost removed the crossover distortion, but never eliminated it. This drove the quiescent current up to 160 mA. Yikes!



Above — A variation of Figure 2 employing diodes instead of an NPN level-shifter. To kill the cross-over distortion, I lowered the 4K7 resistor by a magnitude of 10. This gave a maximum clean power of 766 mW with a quiescent current of nearly 72 mA. Head room and quiescent current are inferior to the Figure 1 circuit.



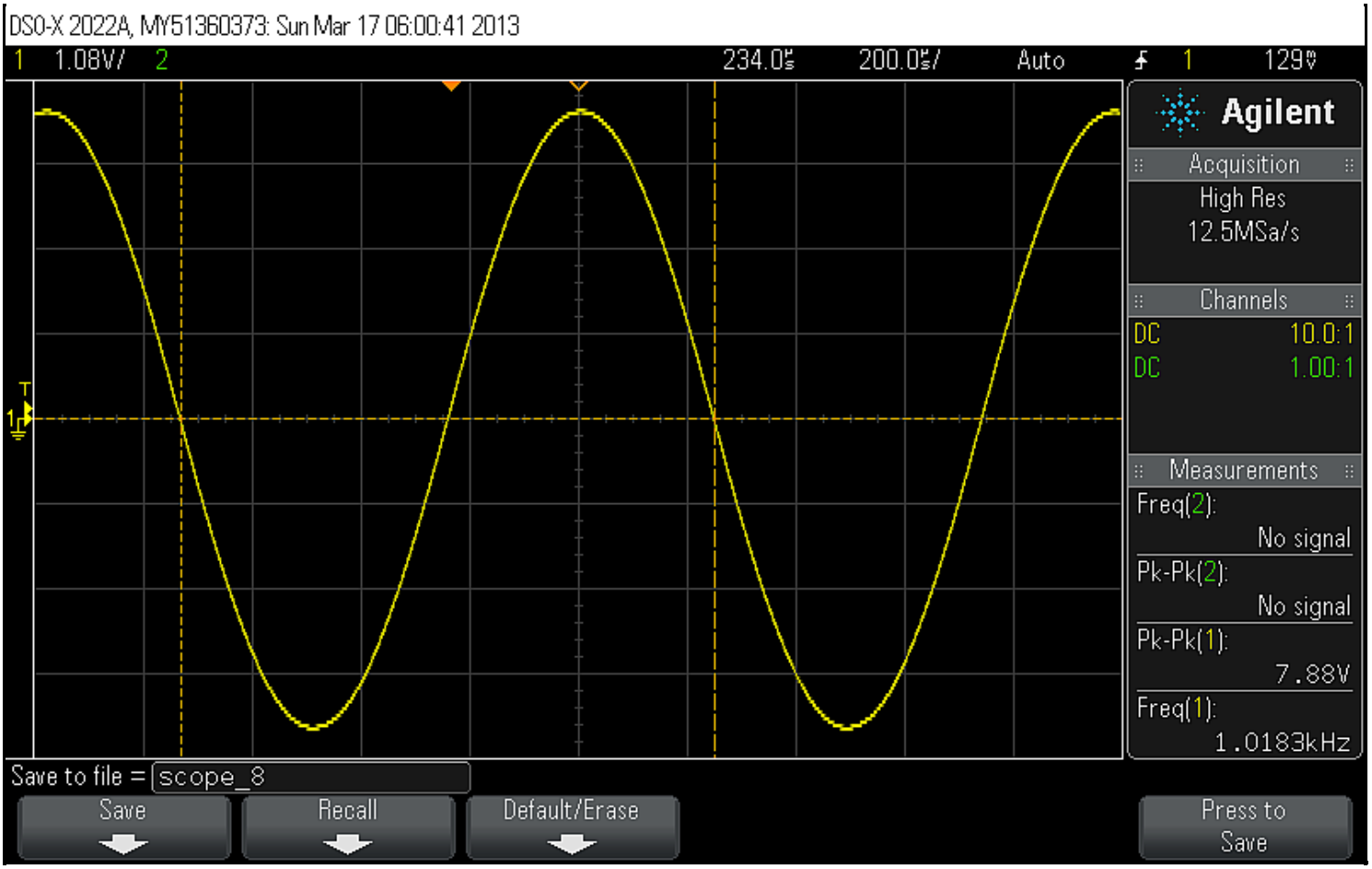
Above — The Figure 3 'scope tracing. Click for a 'scope tracing with the signal generator amplitude increased *slightly* to push this amp into clipping.



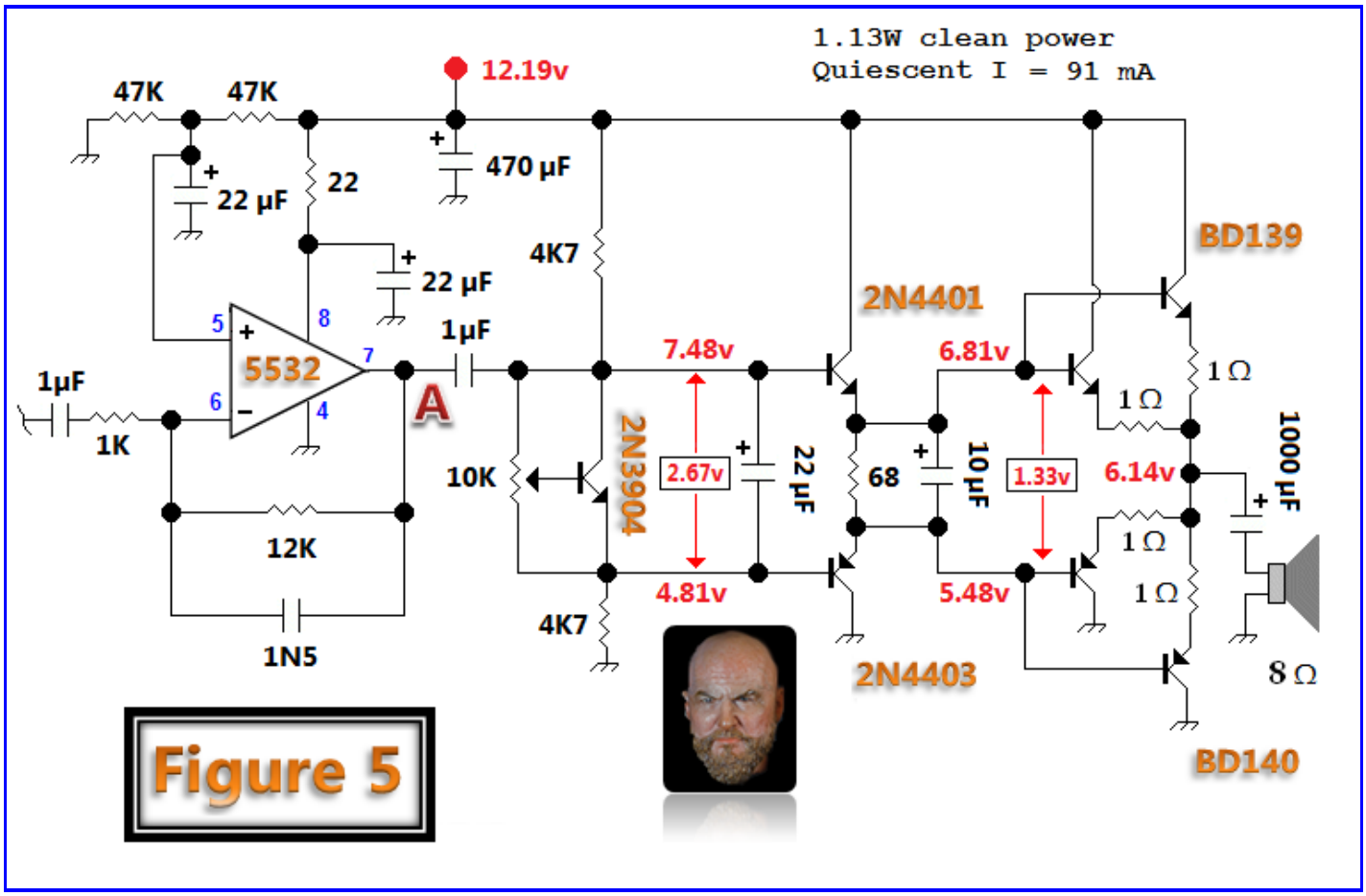
Above — Back to an AC coupled power amplifer like Figure 1. I added a set of intermediate followers built with a 2N4401/2N4403 pair. The clean output power now lies at 970 mW with a quiescent current under 50 mA. Adjusting the trimmer potentiometer on the level-shifter even a tiny amount may change the quiescent current dramatically.

I found a bias of 1.37v across the BD139/140 pair removed all trace of cross-over distortion at maximum clean signal power. Just tweak the 10K trimmer potentiometer while looking at your 'scope and decide what bias you prefer. I lower the bias until crossover distortion appears and slowly tweak it to find the sweet spot. Then measure the DC voltage across the power follower base terminals, plus the total stage quiescent current with the signal generator switched off. You might have to repeat this procedure a few times, since trimmer pot adjustment is quite sensitive.

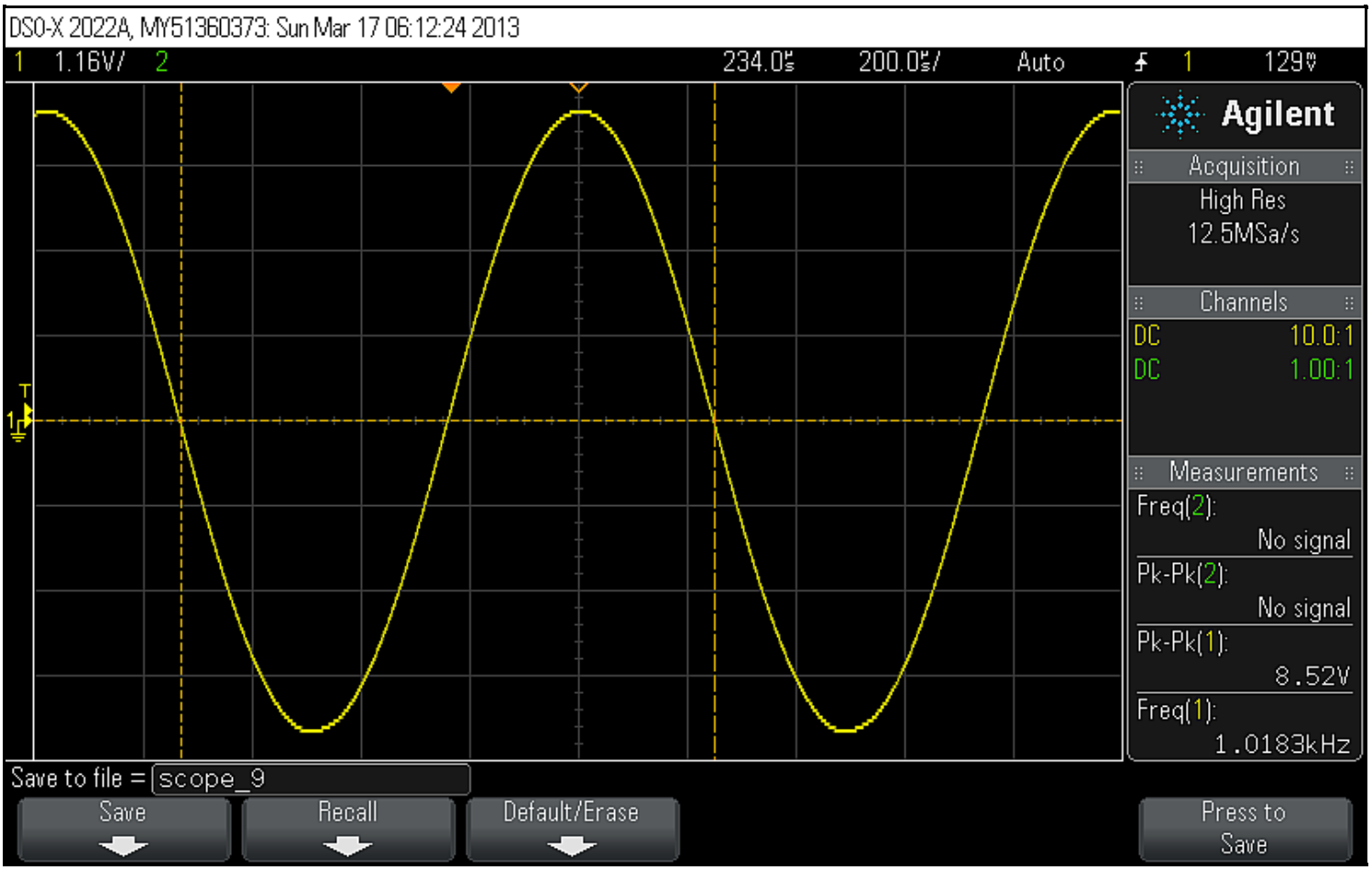
Jerry, W5JH made a PC board. Click for front photo. Click for rear photo.



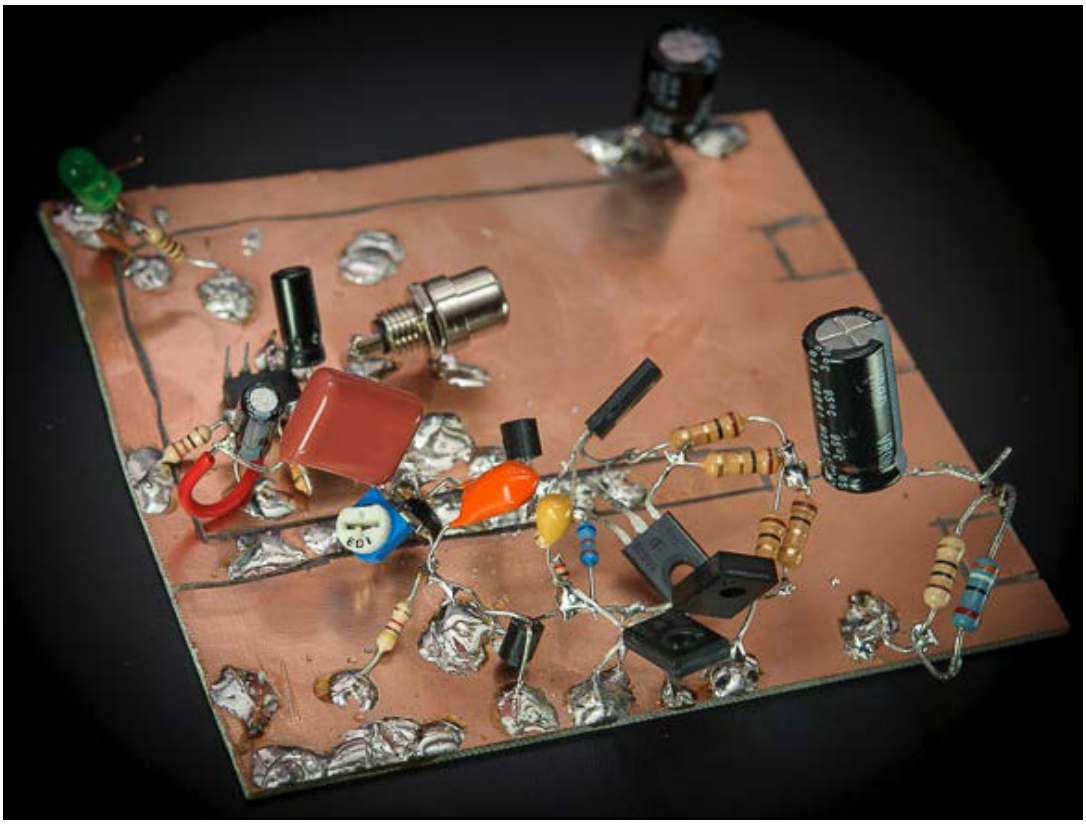
Above — The Figure 4 'scope tracing.



Above — I added another BD139/140 power follower pair in parallel. The boost in headroom over Figure 4 was small, but it was nice to break the 1 Watt barrier. This amp sounded great and blew away an LM386 set up for a gain of 20 — more headroom, less noise and boosted warmth.



Above — The Figure 5 'scope tracing.

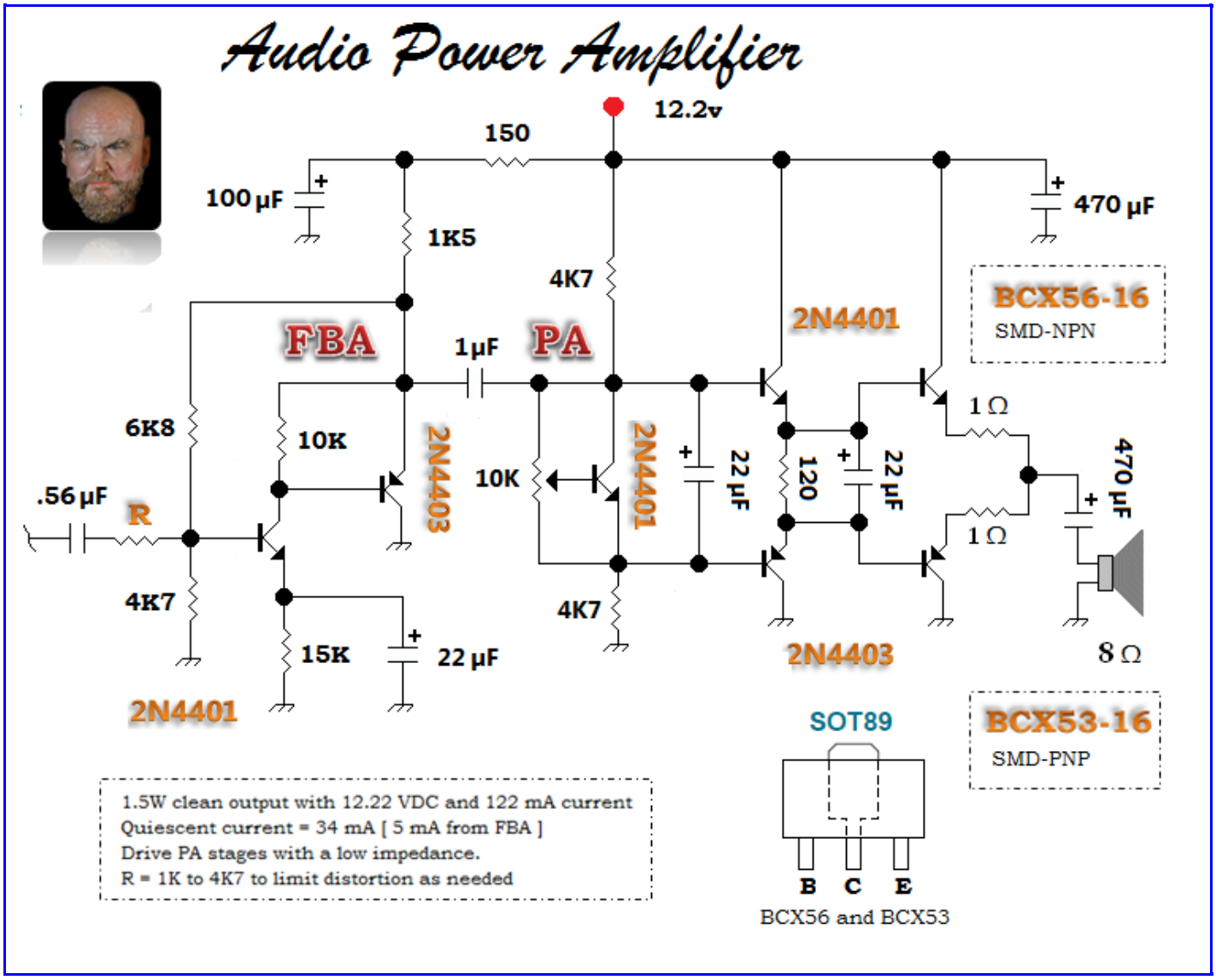


Above — The Figure 5 breadboard. I built all the AF power amps on this board. Signal caps <= to 1 µF were polyester film, while I employed 10 or 22 µF tantalum caps for the level shifters. Electrolytic caps work fine; especially for the level shifter capacitors. The green power indicator LED drew 10 mA and I subtracted this from the quiescent current measurements.

Depending on the AF gain of your receiver, you might wish to add the familiar Zobel filter; a 10 Ω R in series with 0.1 µF C from the positive end of the output capacitor to ground, or more AF bypassing/ decoupling to the circuitry.

No component values were critical — imbuing the spirit of homebrew radio, substitute parts and measure outcomes.

QRP — PosData for August 18, 2014



Above — I built a 100% discrete version of Figure 4 power amplifer with a 2N4401/2N4403 feedback voltage amp that drives the PA stage with a low impedance. Without a low impedance drive, the PA [all the components right of the 1 µF capacitor where you see **PA**] the headroom of the PA section falls down. Thus an op-amp or emitter follower drive works great.

I moved to SMD power transistors: the NXP BCX56 + BCX53. These give a little more output than the BD139/140, and take much less board room. I ran no heat sink other than 8 mm by 4 mm PC board traces for the collectors. A better choice might be the related BCP56 + BCP53 pair in SOT223 since the bigger package of this version better sinks heat.

I determined the optimum emitter resistor for the 2N4401/2N4403 followers = 120 Ω [not 68 Ω as shown in Figure 4]. This drops the the PA quiescent current in half and yields the same performance. If you build Figure 4, change to 120 Ω.

I meant Figure 4 and the QRP POS-Data amp as easy-to-make power amplifiers with a *little* gain. You will have to add voltage gain in some AF projects. For Figure 4, it's easy to change up the 12K resistor, or apply the other half of the 5532. Decouple and bypass better if you get motor boating or hum. 73 de VE7BPO

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9. The Progressive Receiver by Wes, W7ZOI and John, K5IRK

When introduced in QST for November 1981, the Progressive Communications Receiver (PR) by Wes, W7ZOI and John, K5IRK set a dynamic range benchmark for dual-conversion homebrew receivers: 94 dB in CW mode. 31 years later, few other home-built radios have ever reached this benchmark.

The PR is a single conversion superheterodyne 80m receiver; or perhaps it's a direct conversion receiver with an extra mixer ahead of the product detector? It's both and that's the point. Further, Wes and John added other bands with another mixer, plus a crystal oscillator, RF filter +/- an RF amp for each band.

You'll see the PR listed as A High-Performance Communication Receiver in ARRL Handbooks from the early 90s or so. ARRL staffers built and enjoyed the PR in their test lab for many years.

After purchasing the 1991 ARRL Handbook and reading about this project, I slowly adopted and entrenched the PR's progressive (modular) approach. When I built and web published my 2 TRF WWV receivers with crystal IF filters, a couple of people wrote that I "grossly over-designed" them, however, other, more astute builders placed a mixer on the front end and turned their TRF into a multi-band superhet like I later did. These builders understood how progressive circuit building works — all PR inspired.

I hope my introduction renews your interest in the PR. Wes and John's article is thorough and complete — I can't add to it, however, I'll share my thoughts along with those from Wes and John all these decades later.

The IF Stage gain comes from 2 dual gate MOSFET amplifiers. The final MOSFET amp drives a BJT differential pair providing 9 MHz RF to the product detector at 50 Ω and signal for the AGC circuit. Wes recalls using a 3N211 in the original MOSFET slots since Doug DeMaw owned a pile of them and contributed some. In some areas, the 40673 was the dominant 2-gate MOSFET soldered by radio home builders back in the day.

An evolved version of the PR Intermediate Frequency stage appeared in EMRFD as Figure 6.50. Wes replaced the now hard-to-find and *expensive* leaded dual-gate dual gate MOSFET with a cascode of J310s. I built this stage and it performed well with my 12.2 VDC power supply.

Jeff, WA7MLH built the general purpose IF system and noticed if the power supply dropped below 12 VDC, the gain control and maximum gain fell off and he wrote to Wes about his findings. Wes later confirmed dysfunction with a lower DC supply. I asked him to recollect this time:

"…. Frankly I don't remember if I went immediately to computer simulations or if I built a single stage. I think I built. Anyway, Jeff was correct, depending upon the FETs that were used. Then I got to plowing into the details of the cascode, this time with SPICE. I was using PSPICE for some simulations, but was in the process of switching over to LT-SPICE. That makes no difference, for I used the same models in both.

Anyway, playing with a single stage showed that the mechanism for gain reduction was that when the voltage on the upper gate was reduced, it did nothing to the upper stage, but it compressed the supply on the lower FET. So far as signal goes, the upper part was nothing more than a common gate stage that passed whatever signal current was there in the drain of the lower part on into the source of the upper part.

But these were depletion mode FETs, the normal thing for most of the JFETs we use. As such, you have to get the gate down pretty low to get the source voltage low enough to be effective in reducing the gain of that lower FET. That's when it became clear that one could use other parts in the upper slot. An enhancement mode FET such as a MOSFET would work well. And just a common bipolar would do the job nicely too..." *Wes, W7ZOI per an email - January 2013.*

In QST for 2007, a refined version of the EMRFD Figure 6.50 IF stage appeared with a hybrid cascode amplifier instead of the cascode JFETs (and the 2 original dual gate MOSFETs of the PR). Further, Wes added a third amplifier to widen gain control. The hybrid cascade stage has since gone in 100s of receivers across the globe; ensuring the PR legacy lives on as the modernized hycas version.

Roger, KA7EXM sold kits of the hycas amp for years, however, these kits sold out. PC boards may still be available via his web site. Roger received permission to publish the QST hycas article online:

Above — The now *sold out* hycas IF System kit once offered by Roger, KA7EXM.

To underscore my love of the original PR IF amp, I built a version using SMT dual gate MOSFETs in 2013 and feel the SMT version might be perfect for builders who prefer to manually control IF gain in most situations, but want AGC control now and then. Versions built with the BF998 MOSFET may suffer parasitic ocillations at UHF and so meticulous attention to decoupling and bypassing out to UHF and in some cases double-sided copper board may be required to prevent unwanted ringing at UHF.

If you build the entire PR with the BF998, the VFO output runs about 10 dBm and requires attenuation. Again, please consider UHF oscillation precautions.

Wes and others have built both leaded and all-SMT versions of the hycas amplifer.

John, K5IRK Recalls

I wanted John's recollection of the PR design and build and received the following narrative in late March 2013:

"…One night in late 1979 I had telephoned Wes to chat about QRP and building rigs. During the conversation he told me he had a project and related QST article in mind and asked me if I wanted to participate. He was in the process of writing a book (IRFD) and didn't have much free time, so he needed some help. I told him yes. A few weeks later I received the first schematics and began to gather parts.

Recall that the project started as direct conversion receiver. But it then progressed (hence the title of the article) to a superhet. I choose to build my superhet with a SSB filter as I had already built Wes' CW Competition Grade RX from SSD. During the next year he sent me schematics and a few critical parts thru snail mail. (The Internet and e-mail would come later.) I would build the circuits and report the results back to him. As I recall, we only had about three phone conversations throughout the whole project.

The design was an iterative process. For example, the IF circuitry began with a single 40673, but grew to include two stages. Wes bread boarded this in a mono band version of the receiver at his end. An early audio derived AGC system was replaced by the IF derived circuit that appears in the article. The VFO was also designed twice, as were the Front end Filters and the BFO. More is said below about the design procedure.

I designed the circuit board layouts. The traces on my boards were drawn by hand and etched at home. I sent hand drawn sketches of each of the boards to Doug DeMaw; Circuit Boards Specialists (CBS) then created the commercial boards for those folks who wanted boards. I recall that we sent out layouts to any readers for a SASE. All of my boards were designed and built on double sided PC material with the exception of the VFO board and the Audio Filter boards. This can be seen in the photos. The boards eventually sold by CBS were single sided. All layouts and functionality were confirmed prior to publication.

The traditional design process for a homebrew ham receiver in 1980 started with schematic sketches based upon the intuition and experience of the designer. The total circuit would then be built. Measurements were merely things that were done afterward, something to characterize the result. The PR was different. Individual stages were designed, built, and measured. Negative feedback was used in the critical amplifiers to guarantee that the gains were high enough for reasonable noise figure, but low enough to preserve input intercept.

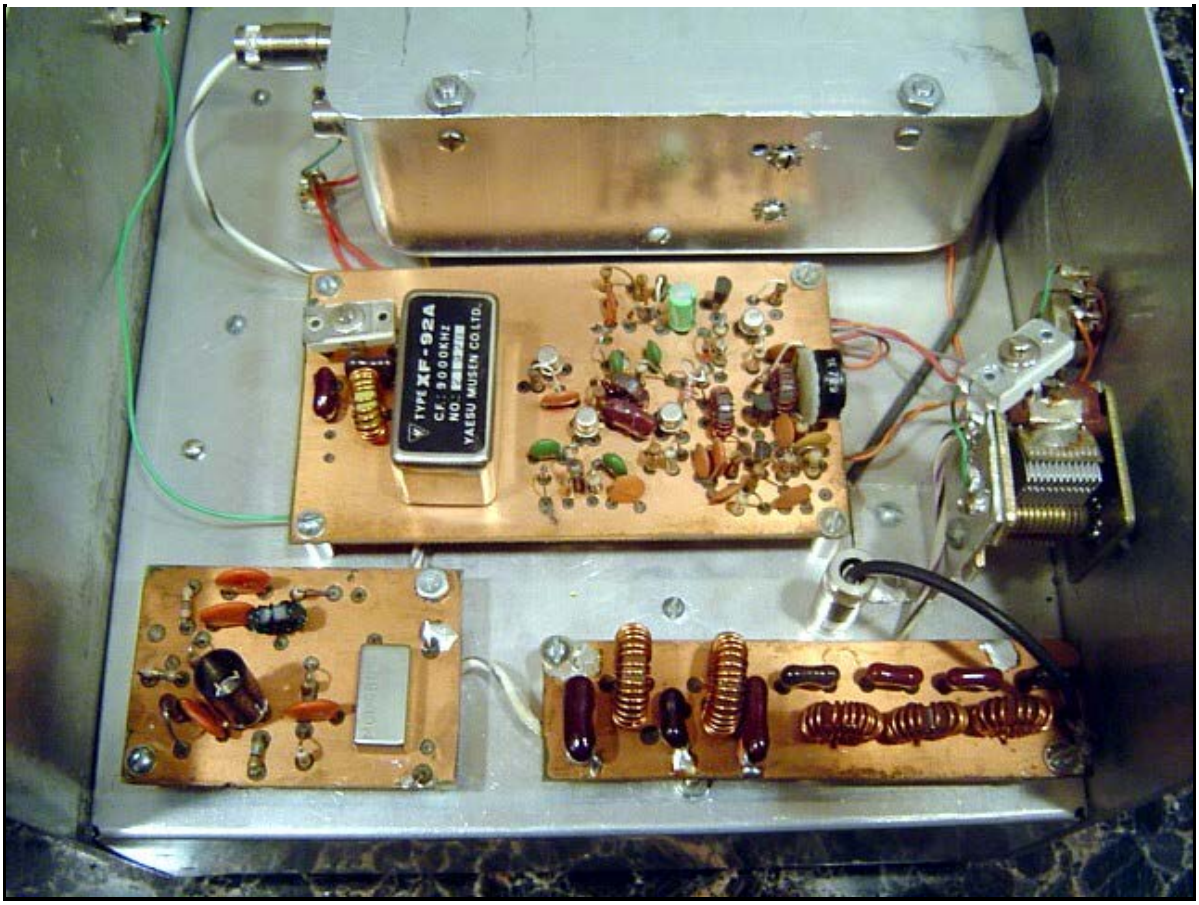
Only after the individual stages were operational, were they assembled to form a working receiver. Some stages were further modified during system assembly and evaluation. We had no computers available for circuit simulation, although the gain distribution was optimized with a hand calculator. The goal was not just a receiver that sounded good, but a box with good two tone dynamic range.

Once my receiver was completed, I sent it to Wes for MDS and DR measurements. A goal was to compare these measurements with those already done in Oregon. Roger (KA7EXM) took many (if not all) of the photos. Wes then forwarded the receiver to ARRL. They returned it to me when they were finished. Wes refined the article during the final months of the project before it was finally published. I received copies of his drafts, and then offered my feedback.

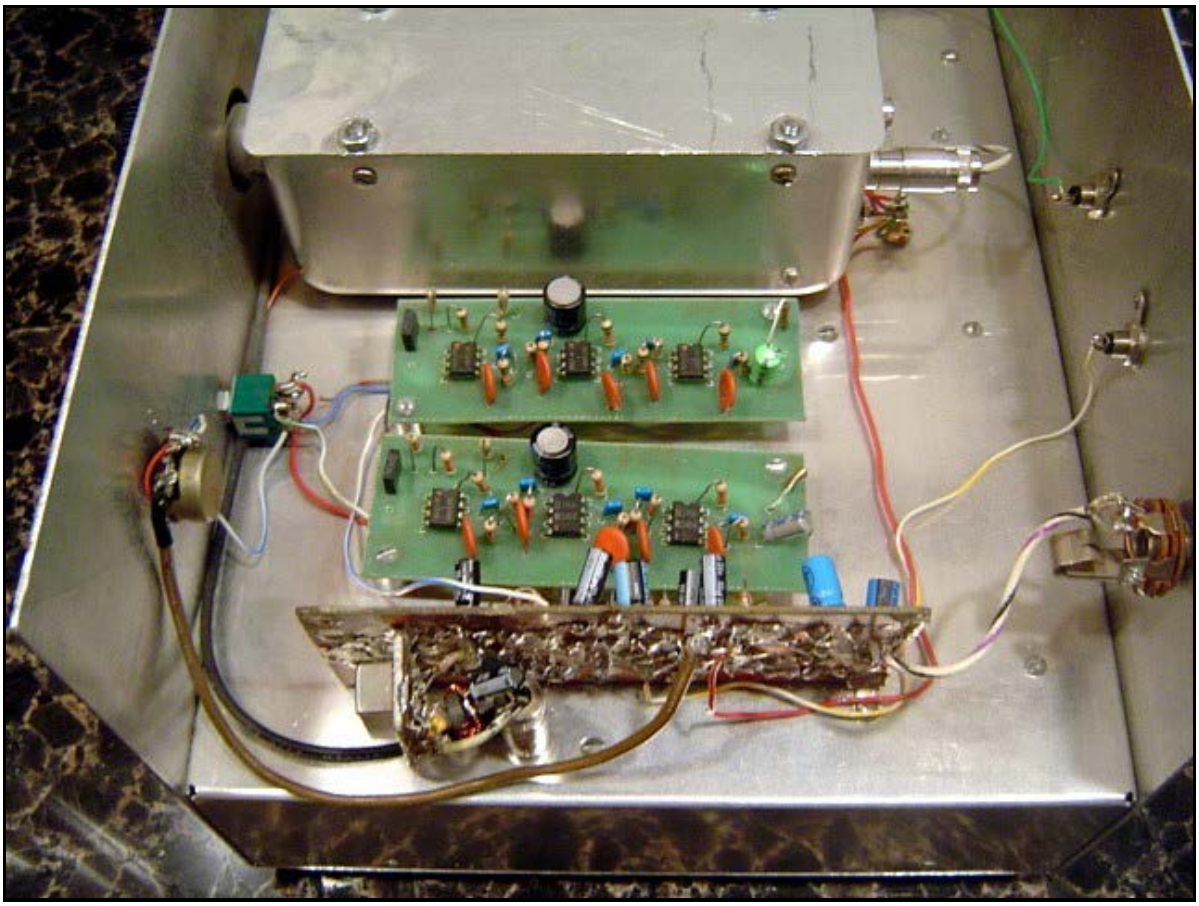
We had hoped that a "few" experimenters would enjoy building the receiver, but had no idea that it would be as successful as it was. We eventually learned why this occurred: First, the receiver held up well when it was measured at ARRL Headquarters, exceeding the performance of most of the appliances being evaluated at the time. This prompted some League staff members to build the receiver for their own use. This, in turn, prompted them to include it in the Handbook for several years.

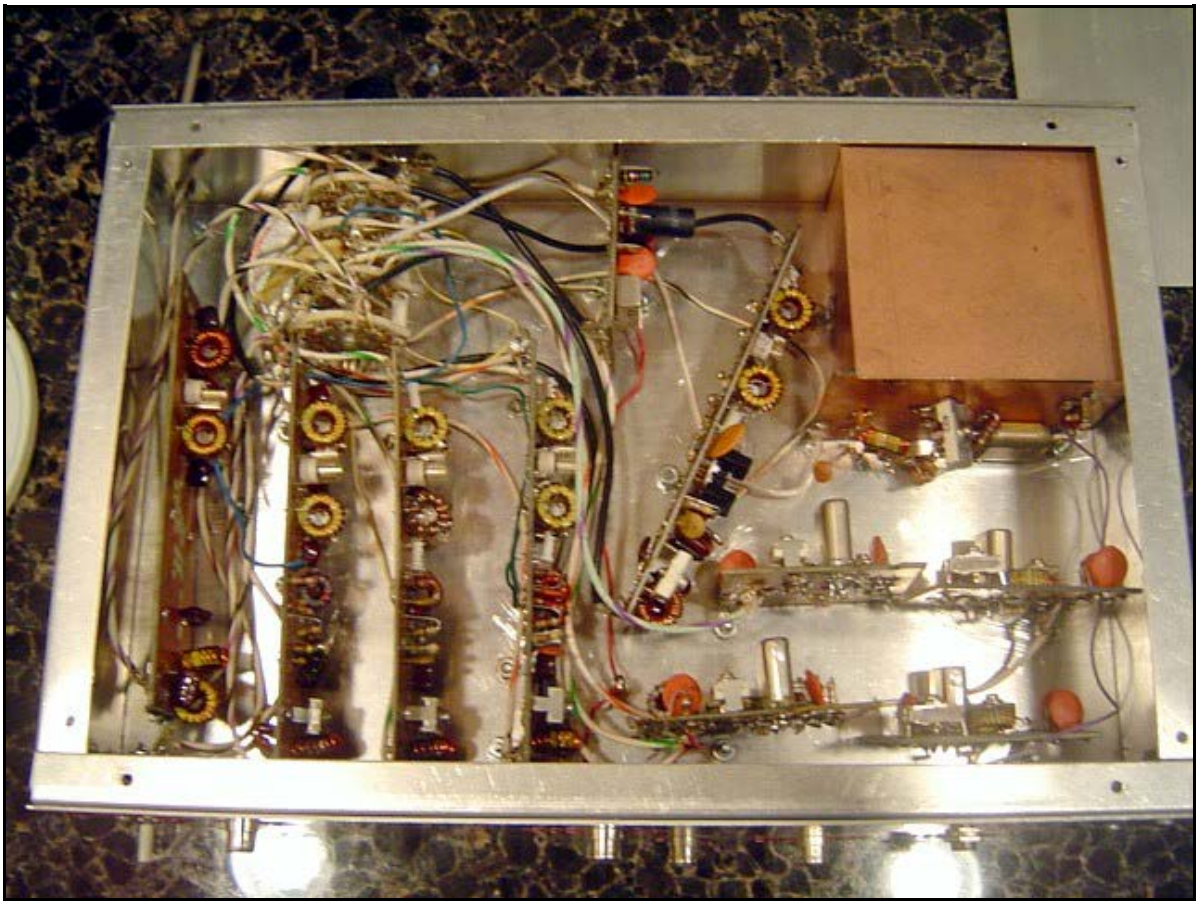
My original receiver plays as well today as it did in the beginning. The electrolytic caps on the audio board have been replaced, for they were beginning to go south on me, but that's the only change. I do have a second version that is used for experiments..."

Some PR photos taken by John in March 2013



"...You will notice a couple of things different in these photos than in the photos in the article. First, tacked on to the Front Panel 365 pF variable cap is a mica trimmer cap.....this is of no significance as I was just messing with lowering the BW upper and lower frequencies to see what difference it would make....None is the answer..." *John, K5IRK.*



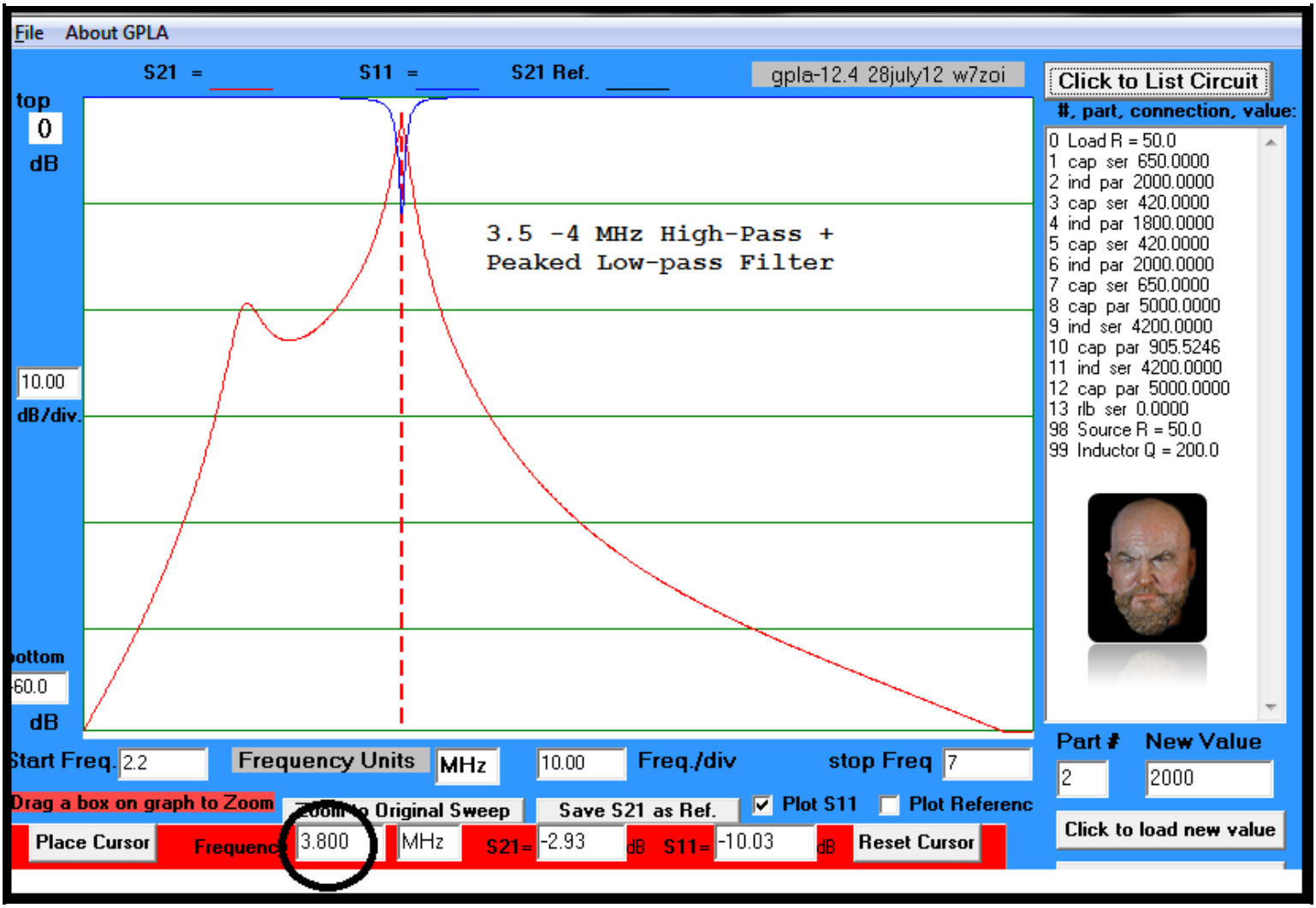


"...In the above photo you will see the addition of the 30 meter band to the RX with the diagonal front end filter board and the Xtal Oscillator attached to the BFO Box's wall that was added some years after the article was published…." *John, K5IRK.*

High-pass plus Peaked Low-pass Filter

If you know me, you know I love the peaked low-pass filter both at RF and AF. This adoration came from studying the PR and other work published and shared privately to me by Wes. *A good reference = The Peaked Lowpass: A Look at the ultraspherical filter by Wes for Ham Radio, June 1984.*

While the high-pass + peaked low-pass filter placed ahead of the second mixer in the PR wouldn't likely go in a modern W7ZOI design, its narrow-band LC filtration generates a roofing action when the receiver uses converters for bands other than 80 meters. The result is that much of the DR obtained in the single conversion version is retained in the dual conversion receivers.



Above — A GPLA simulation of the front end high-pass/peaked low-pass filter centered for 3.8 MHz by tweaking capacitor #10 in the software. Click for another simulated filter that covers 3-4 MHz. Look at the sublime low and high-pass skirt action with a 3 dB bandwidth of 73 KHz. Love this!

Click for a screen shot of an entire 40 Meter band version shown with the low-pass filter section centered at 7.003 MHz. I also made 1 for WWV 10 MHz — by scaling the original version's XL and XC and tweaking with GPLA, you may build 1 for any HF band.

AF Amplifier

The post product detector AF chain sounds great. Many of us later replaced the Q5 mute switch with something quieter, however, even today — this AF block holds its own against most discrete component headphone-level circuits. I particularly love the crisp fidelity of the Class A feedback pair Q3 and Q4. Today, I would substitute a BD139 for Q4.

Wrap up

I could go on, but this web page is already too long. Studying + building stages from the PR; a receiver designed more than 3 decades ago, might raise your game today. Go team!

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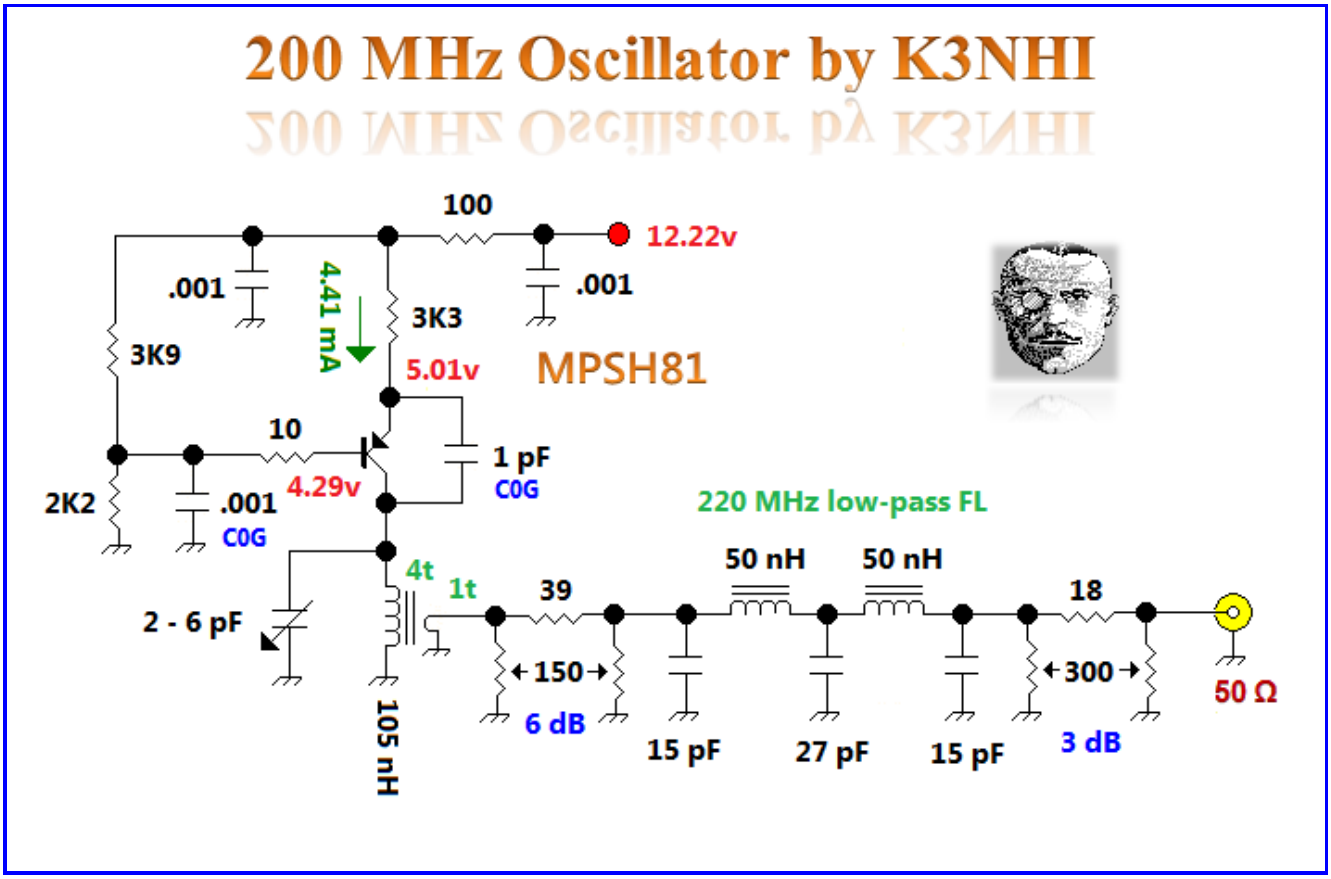
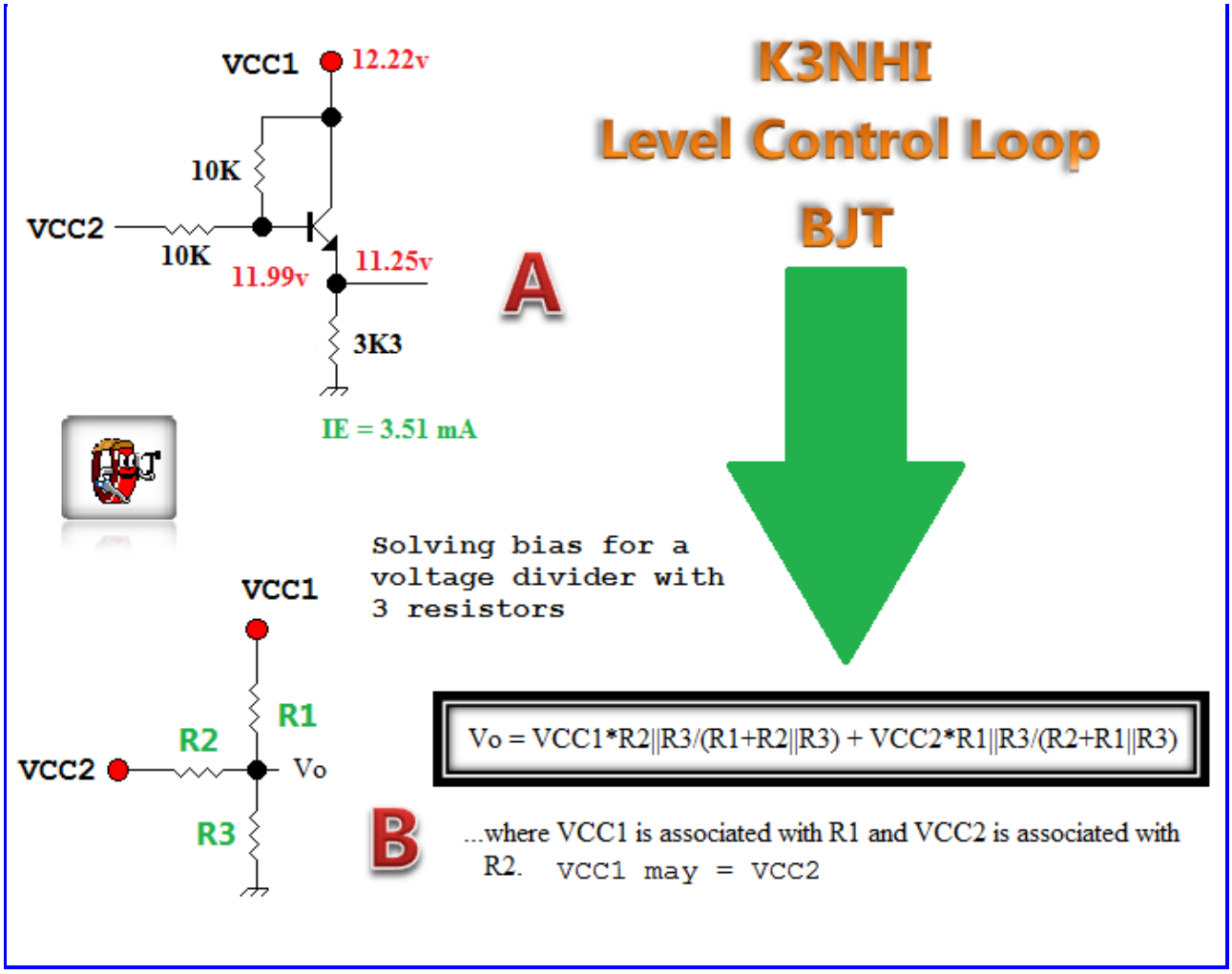
10. Miscellaneous Pictures and Figures

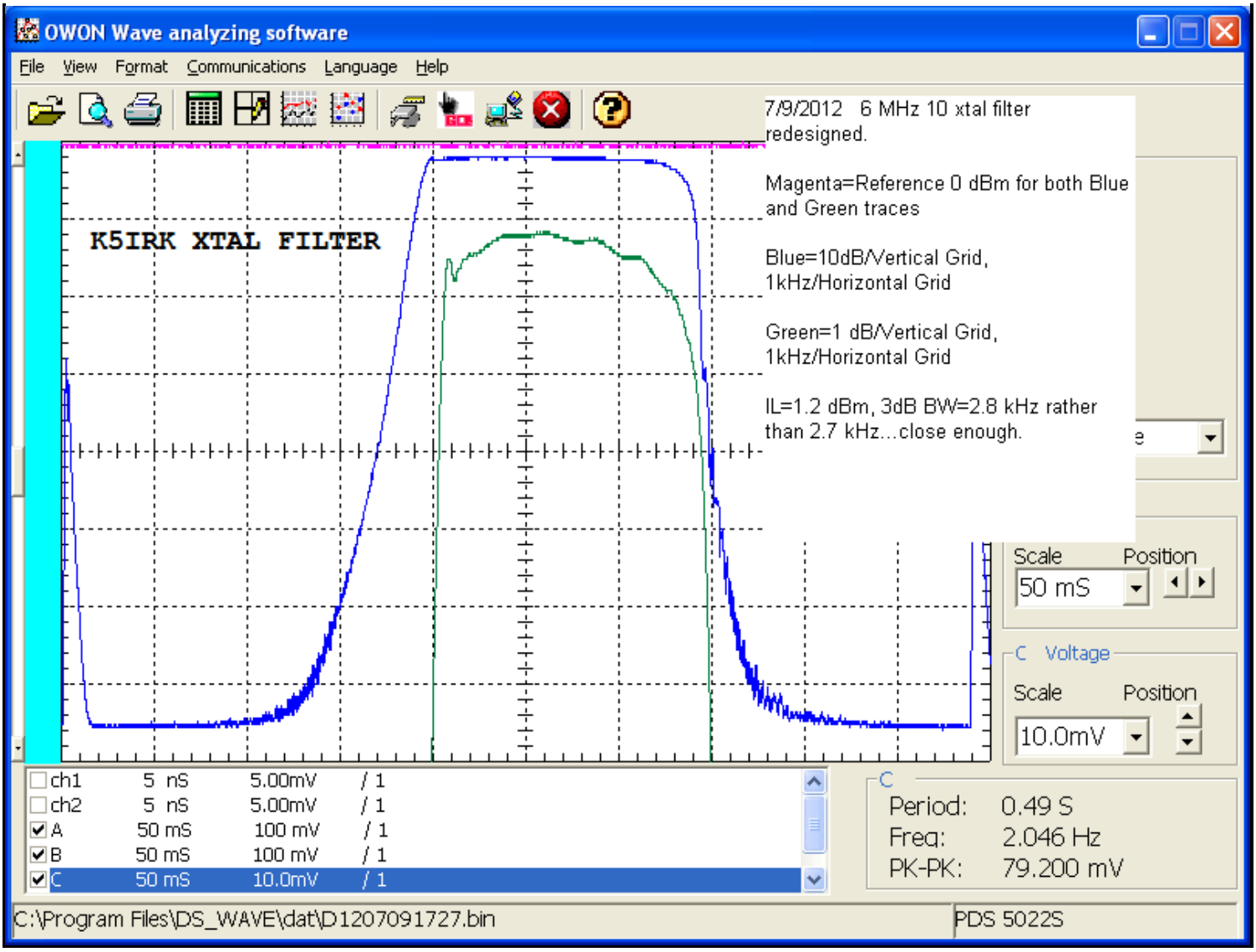
Above — I built a prototype 200 MHz oscillator with trailing low-pass filter before making the 210 MHz version placed in the 1-118 MHz VFO for the K3NHI sweep system. (Section 1). The secondary coil floats and may be positioned between any of the 4 primary links to change coupling and thus output power.



Above — The schematic of my 200 MHz local oscillator. Click for another version I designed that tuned from ~135-208 MHz.



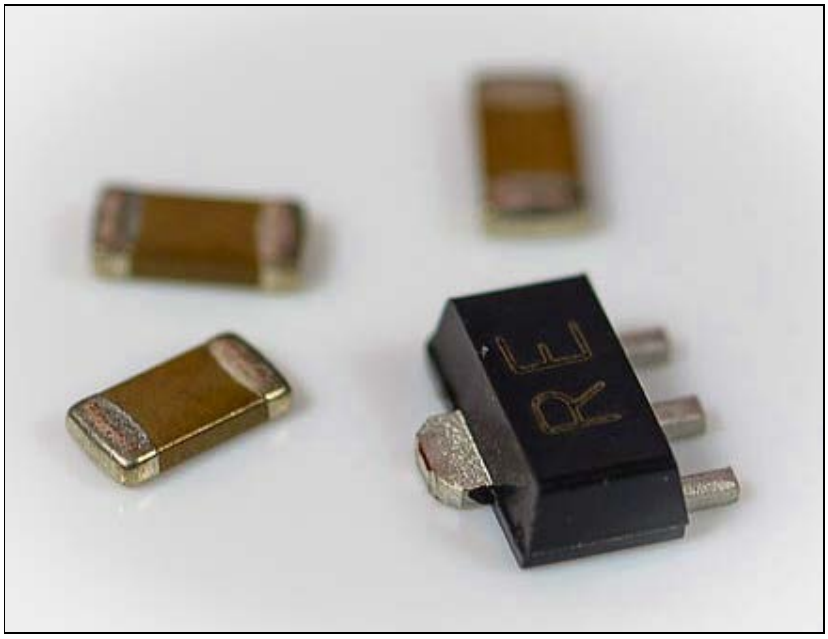
Above — In the K3NHI 1-118 MHZ VCO lies a fabulous leveling circuit that involves a CMOS rail-to-rail op-amp controlling a BJT level shifter. To understand the bias of this BJT, I made a breadboard (A) and then a simple model (B) and developed the equation shown that involves 3 bias resistors. Bias or Vo is determined by superposition. VCC1 and VCC2 may be equal or not. VCC2 comes from the op-amp. I also learned it's a good thing to surround myself with smart people.



Above — John, K5IRQ designed and swept a 10 crystal SSB filter with an insertion loss of 1.2 dB in the K3NHI sweeper that I just had to show you. Click for his build up in Ladbuild08 and then GPLA08 from the EMRFD CD. Bob's xtal sweeper from Section 1 and Wes' software arm us experimenters with solid tools. Nothing can replace measurement and reason.

RF — Test and Measurement

VHF — Véronique



Repository for VHF experiments conducted in 2013 and 2014.

1. Ferrite beads
2. BFS17A transistor
3. 50 Ω MMIC Bench Amplifer
4. Breadboarding Double-Sided Copper Boards
5. Simple Preamp Filter
6. VHF Band-pass Filter Experiments
7. NE612 Mixer Diddy
8. Miscellaneous Photographs

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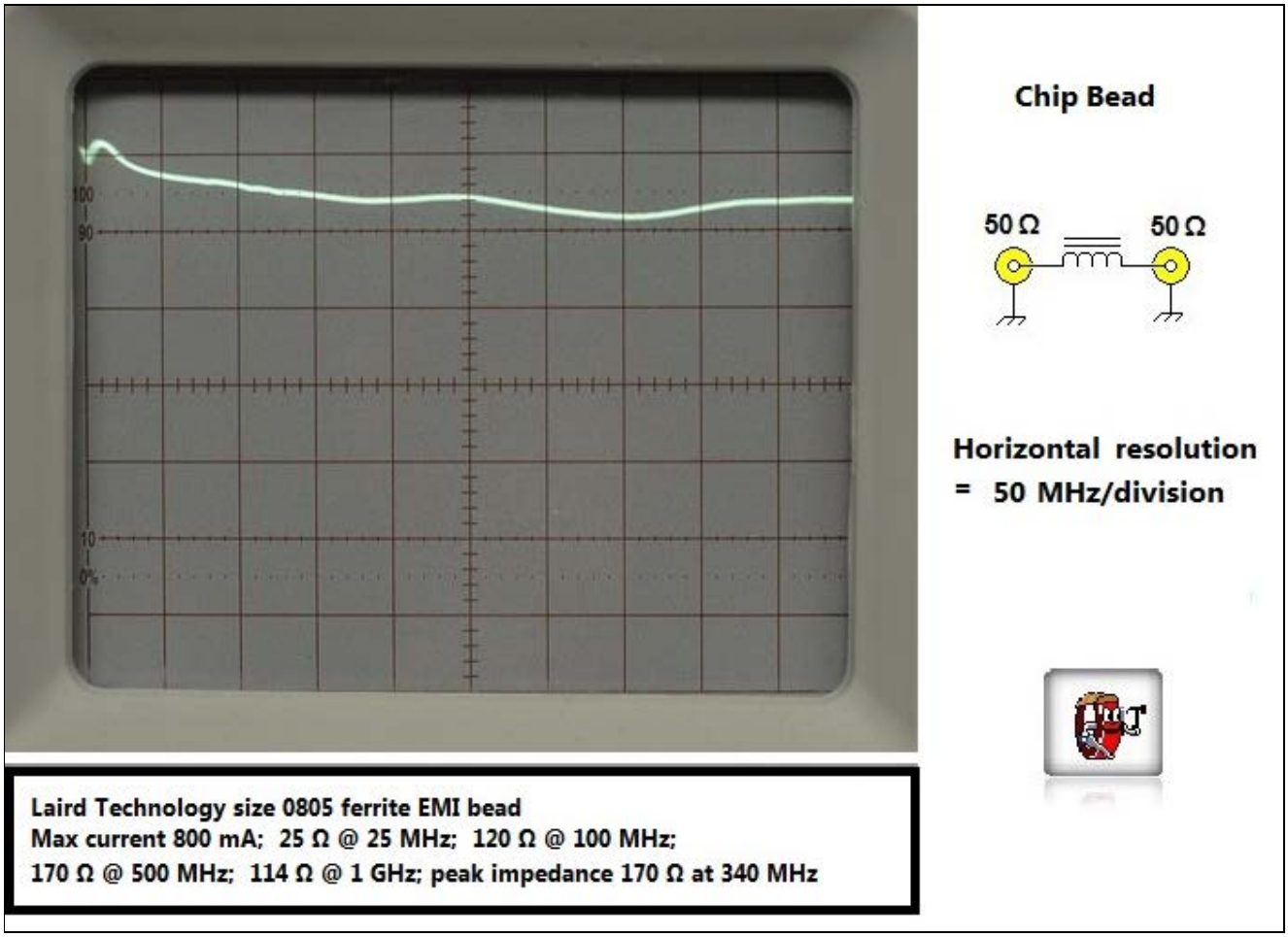
1. Ferrite Beads

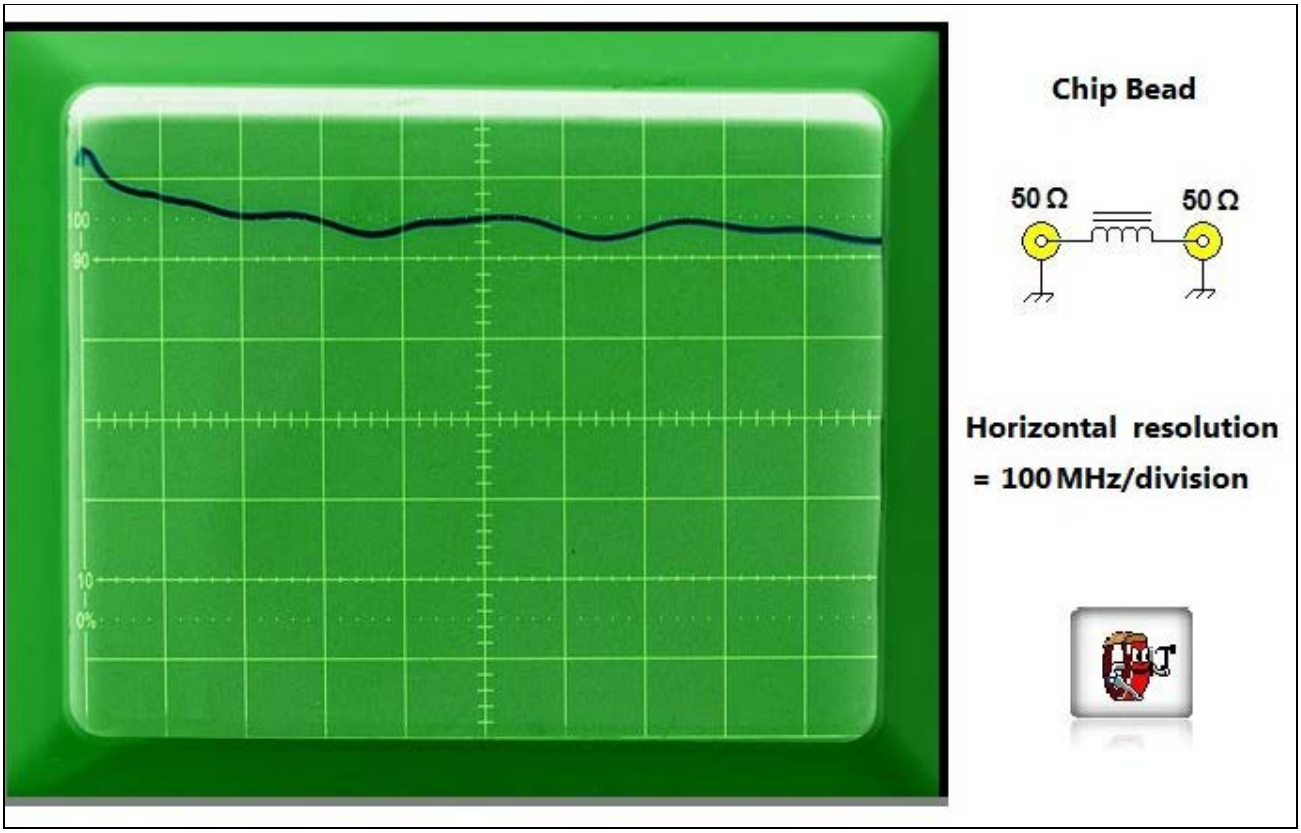
I toiled to choose an SMT ferrite bead for my junk box. Selecting suitable chip beads not only troubles us, but even some engineers I've read — which to choose? Chip beads act as high frequency resistors that present a low impedance to DC plus our desired RF signal while impeding and dissipating UHF through resistive losses caused by minute heating of the ferrite material from eddy currents. SMT beads are typically rated by their maximum current, resistance at DC, and the impedance they present at 100 MHz.

Since a ferrite bead's impedance is essentially resistive to parasitic UHF oscillations, I remembered that the model for a bead is actually an inductor paralleled by a resistor. It's the "resistor" we're after, for that's the extra load that tends to stabilize the amplifier. From past success in UHF supression with low-value (10-51 Ω) resistors, I chose my SMT bead to have a relatively low Z at 100 MHz. Consider, too, that many chip beads sold today serve as RFI suppressors for high speed digital lines where the Z at 100 MHz and maximum current lie well above my requirements.

I found a chip bead that appealed to me on eBay and bought 50: size 0805; 800 mA; with a Z at 100 MHz of 120 Ω.



Above — A sweep of my bead from 1- 500 MHz. Sadly, I did not perfectly center the sweep at 250 MHz, but the peak impedance occurred close to the specified 340 MHz.

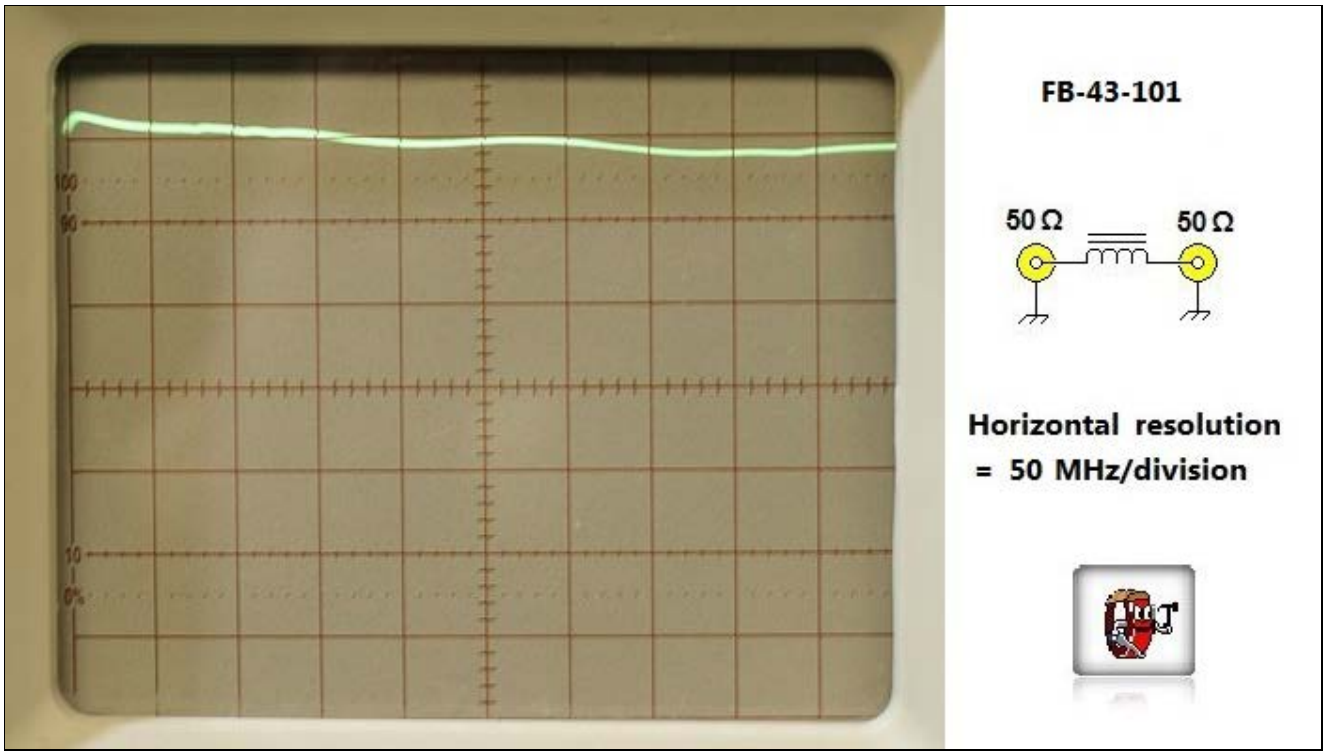


Above — I swept my Laird techlogies chip bead out to 1 GHz. Ultimately, I'll have to verify its function with in-situ experiments. For example, before and after chip bead analysis of a low noise amplifier that's oscillating at UHF.

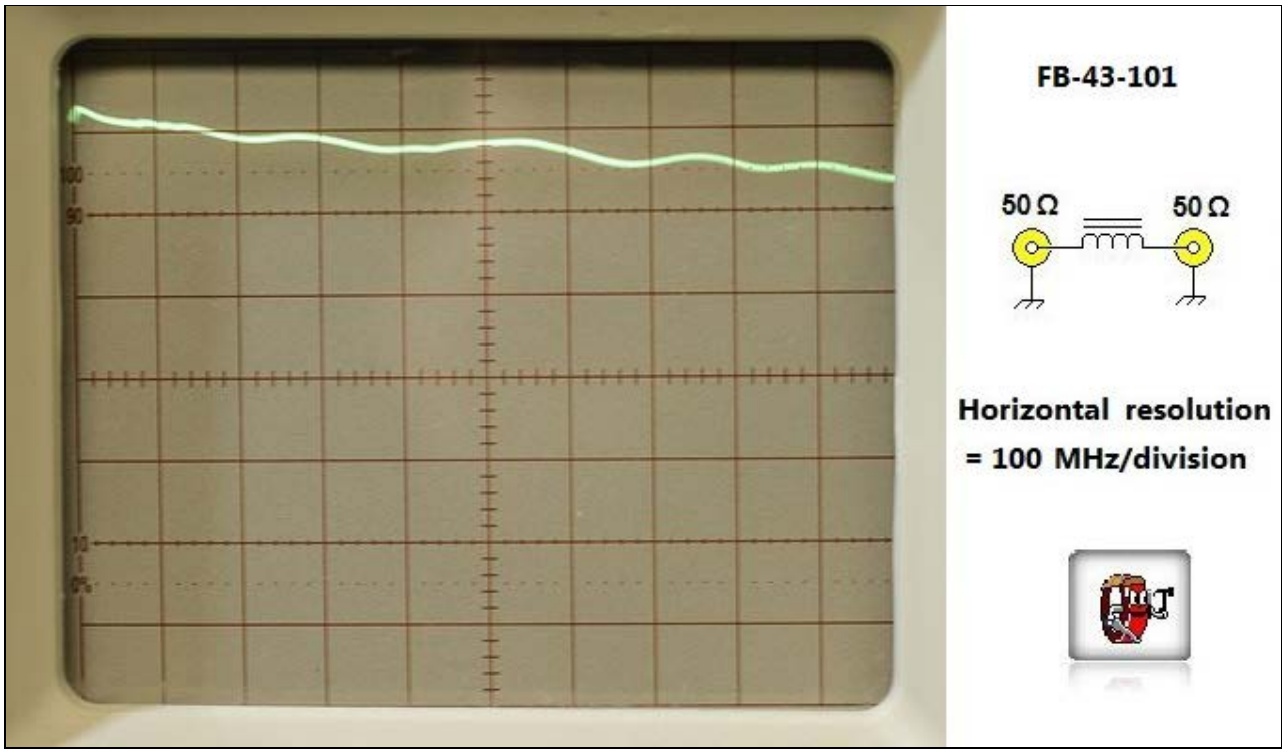
FB43-101

If you sweep the floor under my QRP work bench, you'll find 2N3904s, J310s and a few FB-43-101s in the dust pan. In our hobby, FB-43-101s are common as fleas on a dog; many builders like to stick them on the drain, collector, or base/gate lead(s) of an active part to squash UHF oscillations. I've never seen a datasheet for this part and decided to sweep 1 for reference purposes.

2 sweeps of the FB-43-101 follow:



Above — A 1 - 500 MHz sweep of the FB-43-101 slide-on bead. This tracing shows very subtle attenuation that peaks between 400 and 450 MHz. Probably this minimal attenuation at UHF provides the reason we often need to put 2 on the drain or collector of our amplifiers?

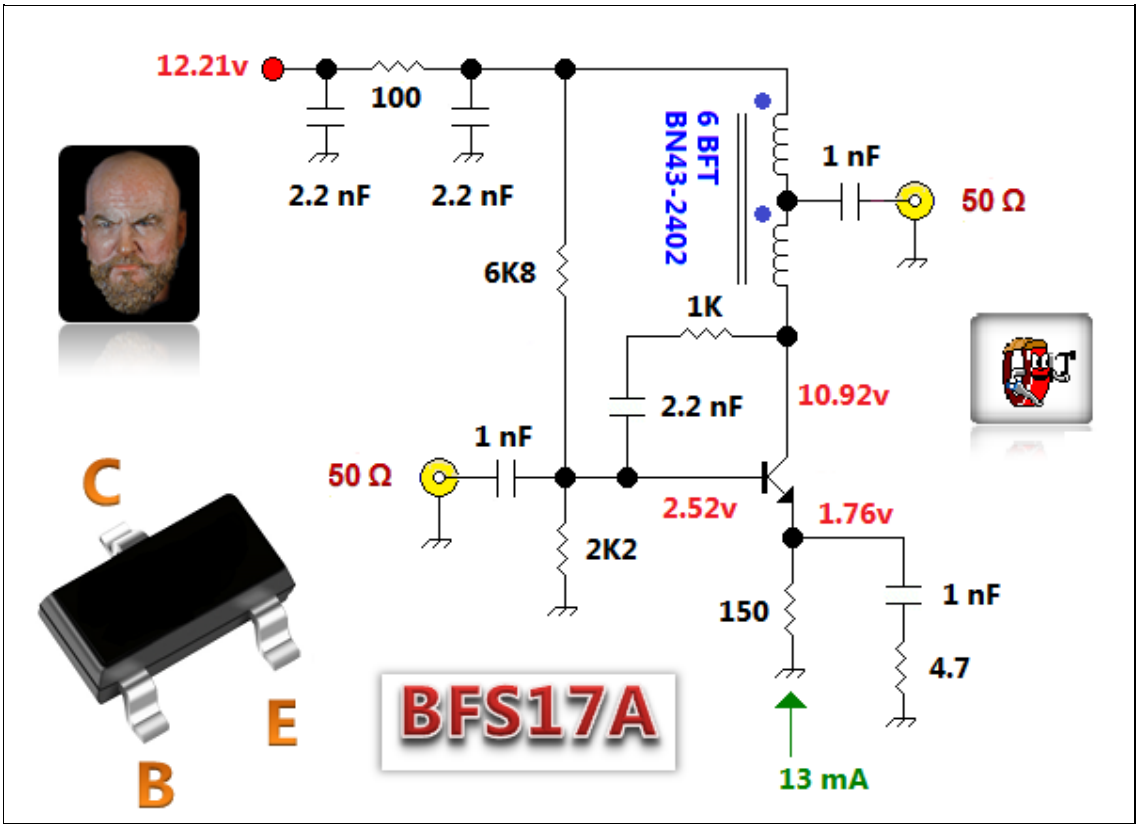


Above — A 1 - 1000 MHz sweep of the FB-43-101.

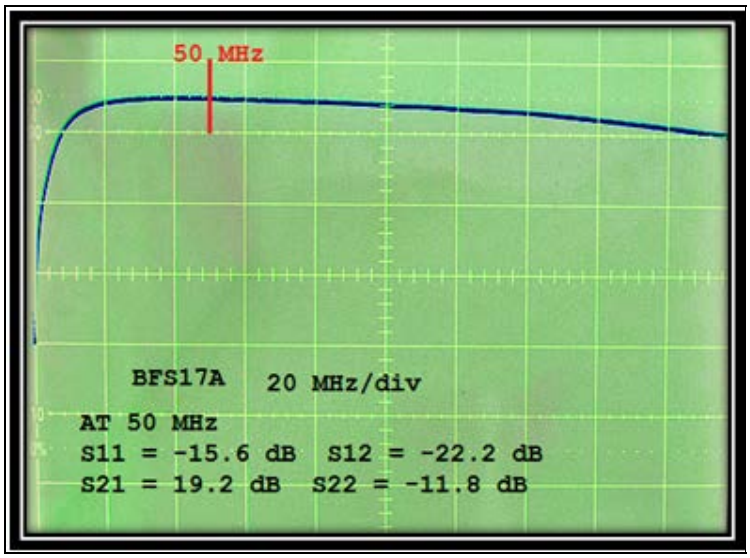
2. BFS17A Transistor

I sought a surface mount NPN transistor to supplant my leaded transistors such as the PN5179, BF199 or MPSH10. Serving as a general purpose transistors biased for ~5 to 20 mA collector current — they'll work as buffers + oscillators + as the BJT for hybrid-cascode amplifiers built with a SMT J310 on the bottom.

My requirements were SOT-23 (3 well *seperated* leads allow easily carved islands in a copper board with a dremel tool); an fT that's not too high to help reduce higher UHF oscillations; and a noise figure < 3. Other factors included price and availability. We enjoy many choices with such criteria, but I settled on the BFS17A. Click for the data sheet . In summary, it's a 2.8 GHz part with a NF of 2.5 dB at 800 MHz.



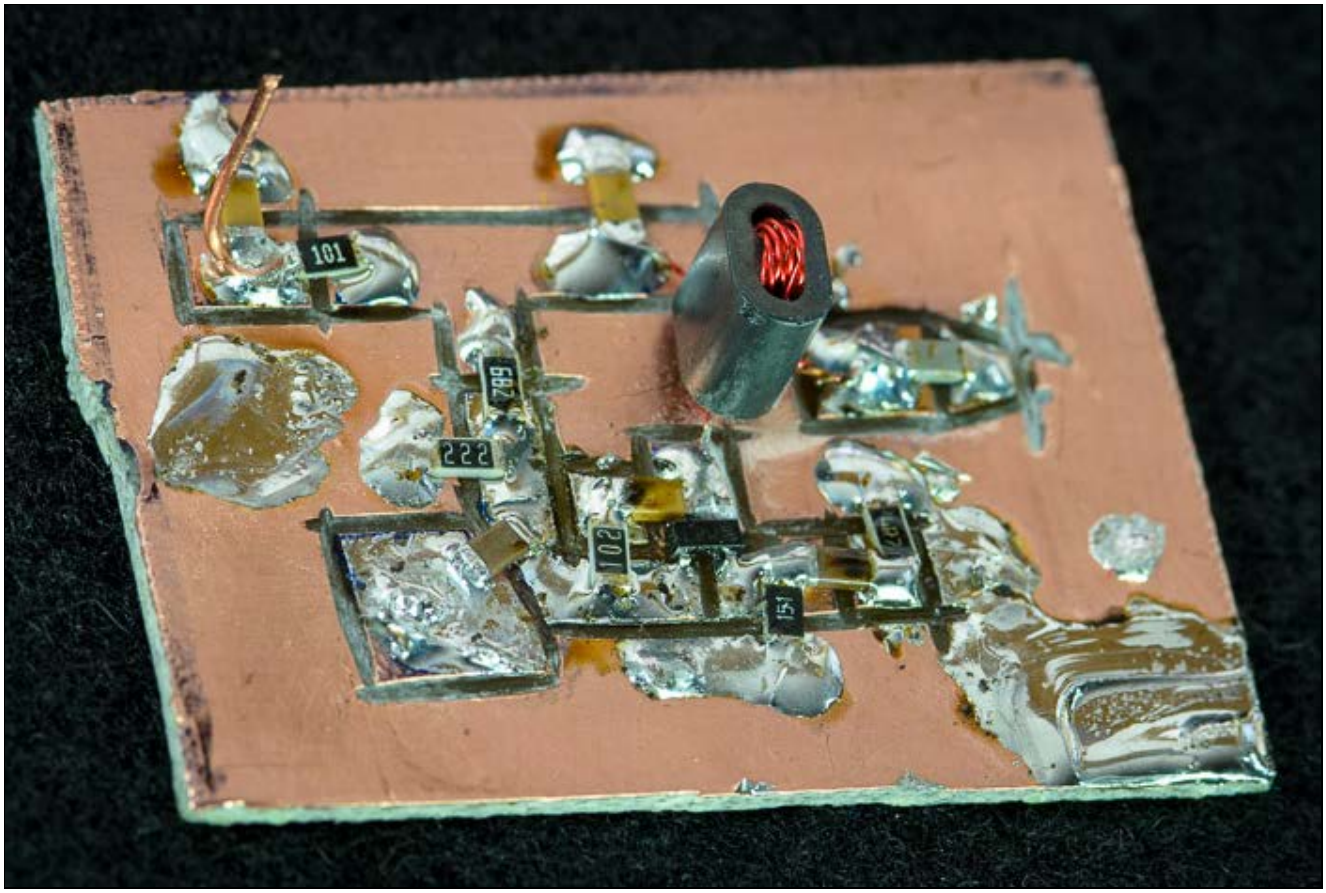
AAbove — My test schematic; a simple feedback amp with capacitors oriented to 50 MHz and above. A limited selection of SMT resistors constrained my experients, but the results seem okay. I strove for the best possible input and output return loss at 50 MHz. I wanted a emitter current of ~ 12 mA and got 13 mA with the bias resistors on-hand.



Above — A 200 MHz wide sweep of my breadboard with tracking generator + spectrum analyzer. Please view the various 2-port measurements

at 50 MHz in the lower half. Despite some input and output port mismatching, a gain of 19.2 dB was measured at 50 MHz.

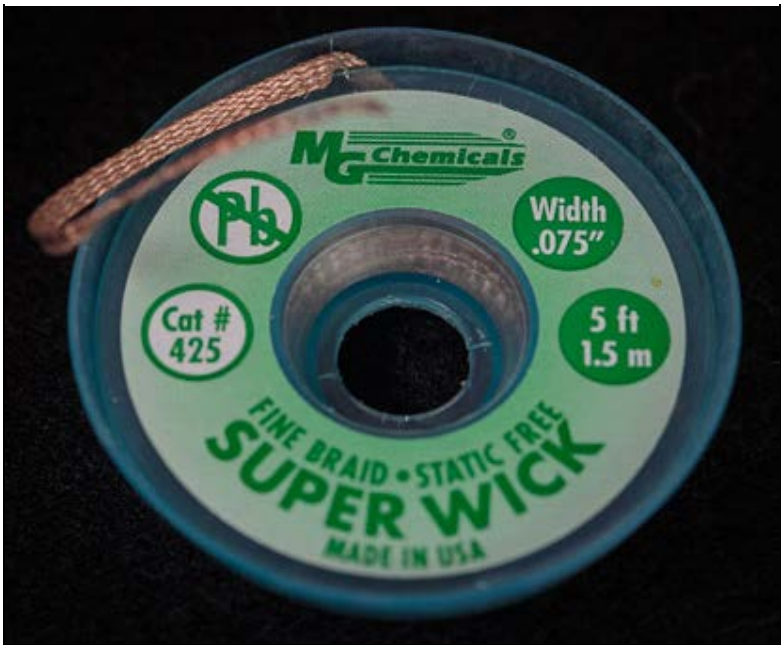
I wound my output transformer within a BN43-2402 binocular ferrite, but ordered some #61 material equivalents for assessment at VHF.



Above — My magnified breadboard looking messy after many parts substitutions from experiments to find the best possible S11 and S22. I removed the shiny RF connectors prior to photography since they often create blown highlights. Alternate image. Despite a limited RF bypass before and after the 100 Ω decoupling resistor + crude bread boarding, I measured no parasitic oscillations from 0.001 to 1 GHz — likely due to the heavy shunt feedback. I liked this transistor, found it easy to breadboard and ordered 100. You'll see the BFS17A in my experiments hereafter.

By all means, order whatever transistor *you* want. Past emailers wrote to say they enjoy reading how other experimenters think about and assess their parts such as RF beads or transistors.

Surface Mount Part Soldering



Above — Solder wick. On my SMT workbench, wick proves the most important tool period. A piece of desoldering wick can transform a monster mistake into a perfect circuit in under 2 seconds.

I've read many accounts of how builders anchor the parts they're soldering. We need to keep the SMT part flat on the board to avoid mechanical tension and poor solder joints — a system is required. For size 0805 and 1206 parts, I often just use my index finger fingernail to hold the part in place while I solder 1 lead with my other hand. After a few seconds of cooling, I solder the other lead and then usually touch up the first joint. For tricky formats like SOT-23, or SOT-143, I usually tape down the part while ensuring even pressure on all leads and just solder 1 lead. After cooling, I remove the tape and easily solder the other leads. Again, the intial lead may require some retouching.

SMT Size

Above — A 5X magnified roll of 50 BFS17A in SOT-23. Compact or what?

Above — BFS17A in SOT-23

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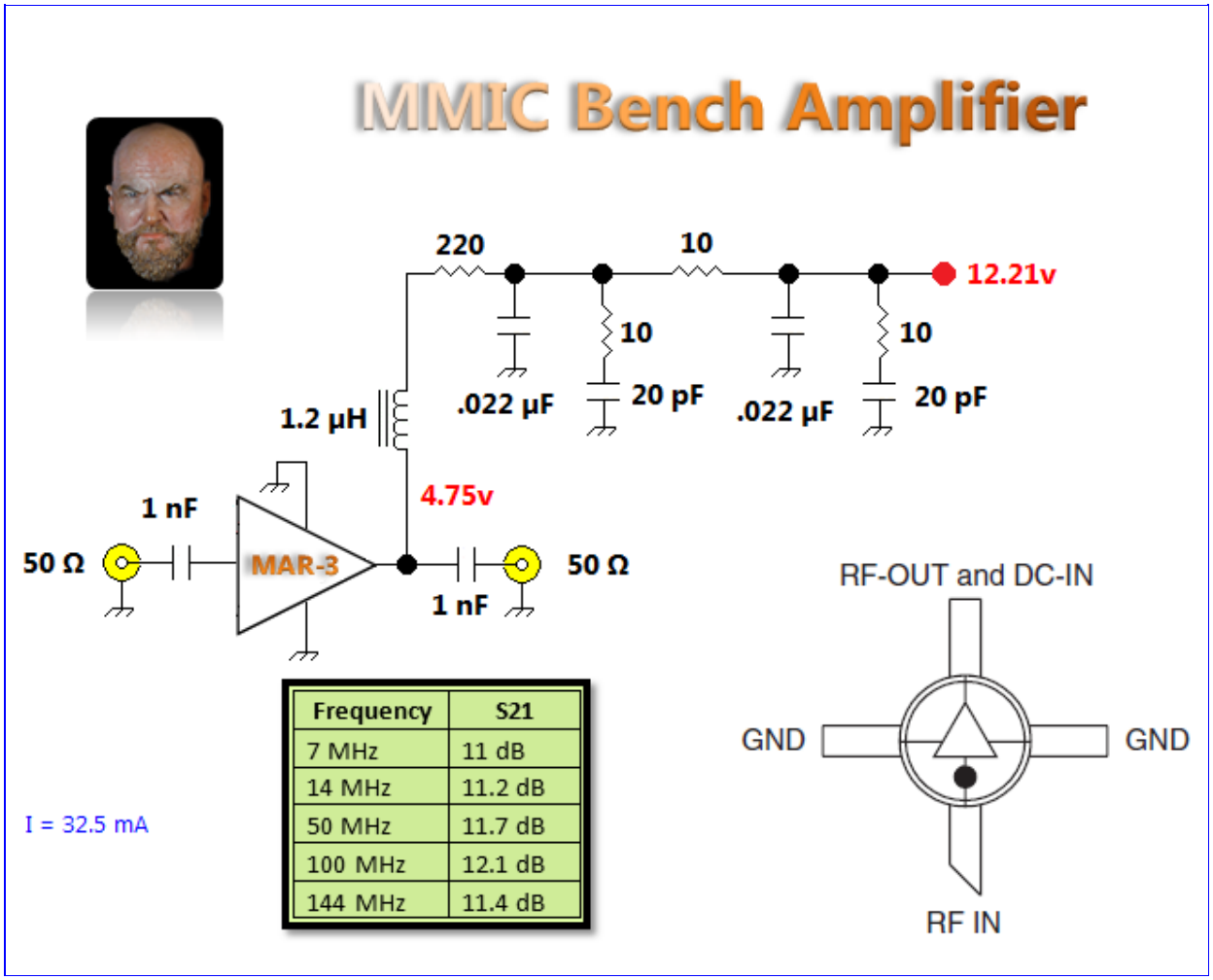
3. 50 Ω MMIC Bench Amplifer

To boost low-level signals during experiments, every 50 Ω workbench needs a utility RF amp module. With strong wideband gain + RF port return loss (S11 and S22), MMICs make a good choice. Since I own 15 pieces, I built an amp around the MCL MAR-3, a modest gain + noise figure, 2 GHz plastic part in the Micro-X package. We enjoy a bevy of MMICS to choose from and as they increase in fT and gain, so does the need for solid, low inductance grounding techniques.

Above — The completed amplifer module in a Hammond box with an RCA jack for the DC voltage.



Frequency	S21
7 MHz	11 dB
14 MHz	11.2 dB
50 MHz	11.7 dB
100 MHz	12.1 dB
144 MHz	11.4 dB

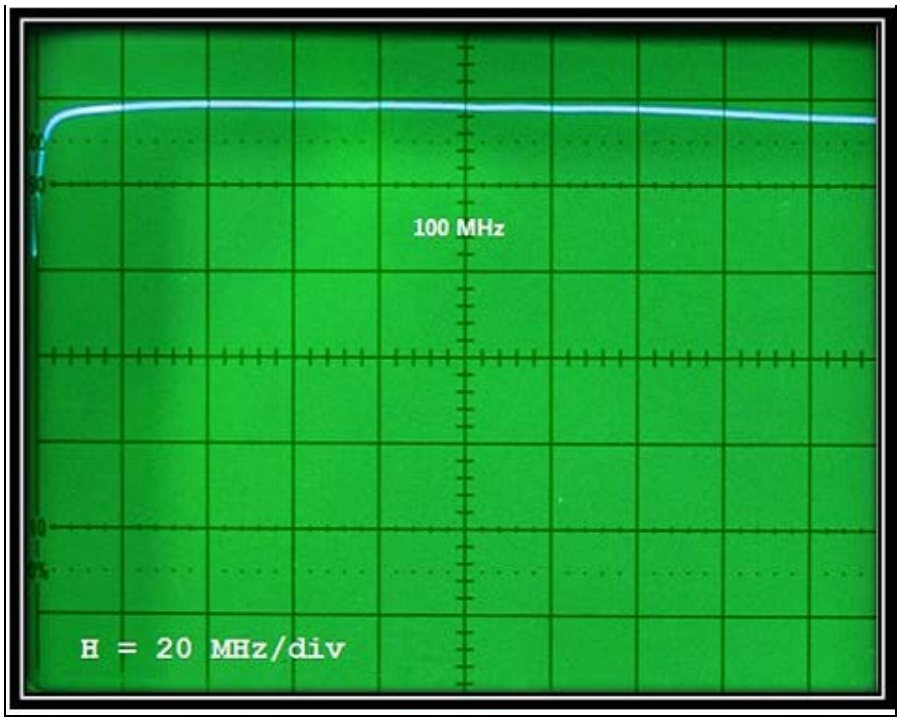
Above — Schematic with a table showing gain versus frequency. I built all SMT (size 0805 mostly) except for *short* leaded 1 nF input and output coupling capacitors. I applied sturdy wideband DC filtering since I can't risk RF flowing on my DC lines during bench experiments. I measured no oscillations from 100 KHz to 1.5 GHz.

I chose a 1.2 µH SMT choke in an attempt to boost gain from ~66 to 150 MHz. Many builders leave off the choke, but some RF will flow through the 220 Ω bias resistor to ground and thus, to get maximum gain, a high impedance RFC might help. Recall this RFC should exert an inductive reactance of at least 500 Ω (>= 10 X the load impedance) at the lowest amplifer operating frequency; in my case this occurs at ~66 MHz. A better choice for the RFC might be 1.6 µH which exerts ~ 500 Ω at 50 MHz, but I just own size 0805 1.0 and 1.2 µH chip inductors.

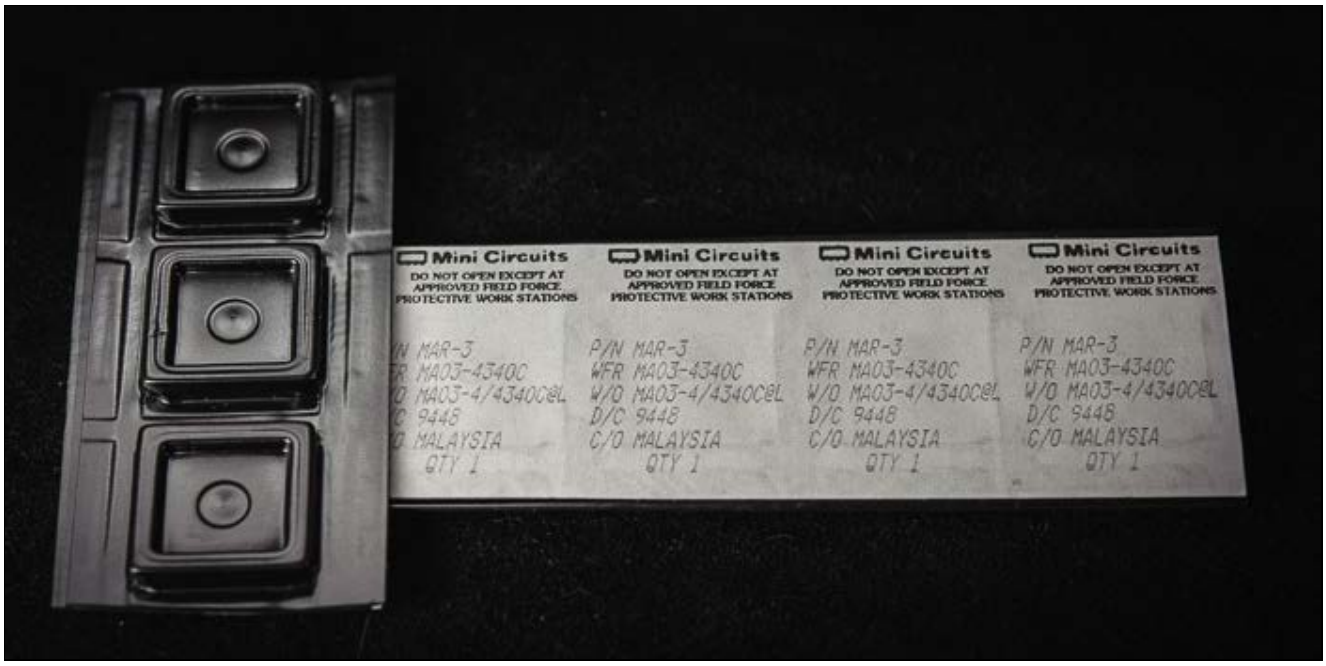
My particular amp delivers ~10 to 12 dB gain from about 5 to 180 MHz. Since the coupling capacitor value determines the low frequency response, if you want a MMIC amp for mostly HF, try reducing your input coupling caps from 0.001 to 0.01 µF or so. You mighty increase the RFC to say 10 µH as well — that's why we don't just build kits; design to suit *your* needs.

The 220 Ω bias resistor should really be 200 Ω for a VCC of 12.2v, however, I had to substitute the 220 Ω R due to a low selection of SMT parts. This dropped the current and also my gain at 100 MHz = 0.4 dB less than that specified in the MAR-3 datasheet — no big deal.

I built the bread board on 2-sided FR4 copper clad board. I soldered thin copper foil around all 4 edges to adjoin the top and bottom copper surfaces. Around each MMIC ground lead, I drilled 4 via holes and soldered copper wire from top to bottom. 2 vias were positioned near every bypass capacitor and all along the input and output paths. 23 total via wires.



Above — Tracking generator plus spectrum analyzer sweep out to 200 MHz. Click for a sweep from 1 to 10 MHz showing a rapid fall off below 5 MHz. This little amp will serve me well for most HF and especially my VHF experiments. A UHF amplifier module is planned and all knowledge gained from VHF circuit building will flow forward.

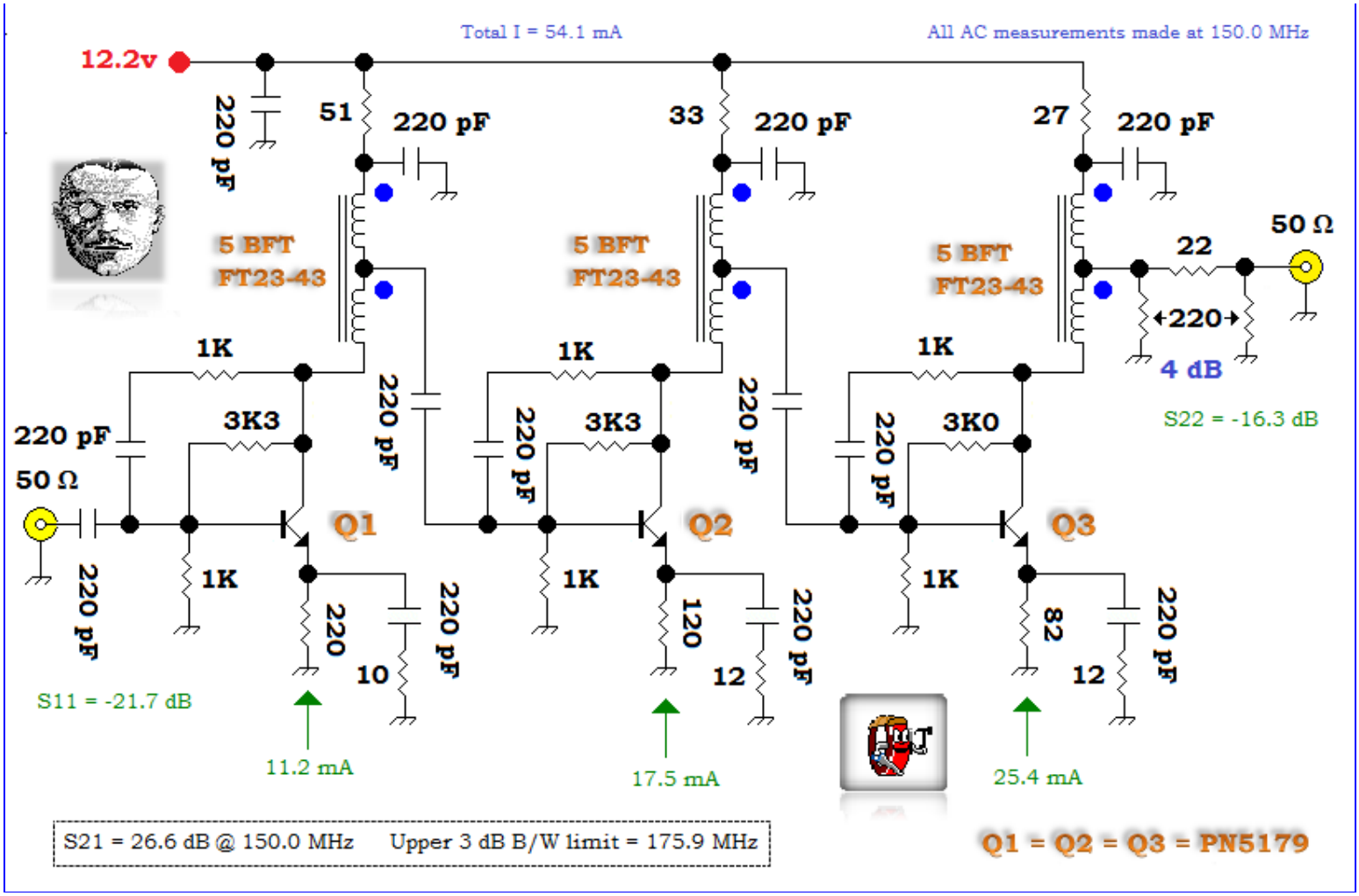


Above — Part of a strip of MCL MAR-3 amplifiers.

Feedback Amps (FBAs)

MMICs like the MAR-3 use a Darlington feedback pair. Still, too, we shouldn't write-off discrete component FBAs wielding transformers + shunt and series feedback at VHF. These amps; staples of countless W7ZOI and W1FB projects since the 1970s; evolved commercially into GaAs FBAs built on a tiny wafer by companies like TriQuint Semiconductor.

In broadband amplifiers, negative feedback permits a wideband (flat) gain response, reduces input and output VSWR (S21 and S22) and controls performance changes from S-parameter and other variations from transistor to transistor. In 2013, I needed an amp chain with >=25 dB gain at 144-165 MHz [~150 MHz mostly] for low level stuff : -25 to -10 dBm input power. Inspired by Wes' 144 MHz CW/DSB transmitter FBA chain *(**SSD** -- Chapter 8 -- Figure 30 in Solid Stage Design for the Radio Amateur; ARRL published in 1986 and out of print)*; I cascaded 3 PN5179 BJTs as FBAs:



Above — My 150 MHz (design frequency) FBA module. Since the fT of the PN5179 = 900 to 2000 MHz @ a 100 MHz test frequency, gain drops steadily as we move up above 100 MHz. Centered at 150 MHz, my upper 3 dB drop off point was 175.9 MHz — thus this amp works okay from 144-165 MHz where I need it.

I built and tweaked each stage to derive the best S11 and S22 . *Click for a snapshot taken of my test of Q1 with a BNC connector attached to both ports — I did this for each stage.* It's a trade off since if you boost the emitter degeneration R, the S11 will improve but the gain decreases. Typical stage S11 = -18.5 dB and S22 = -17 dB. I progressively bumped up the current in each stage to reduce gain compression. I get ~12-13 dBm power out of the whole amplifer before compression/distortion occurs.

Securing the best S11 and S22 for each stages might seem silly, but invoked learning. The S11 and S22 of the *total amplifer* counts the most. An interesting thing happened to S22 – with a RF connector soldered right on the board without the pad, S22 = -23.8 dB. When I added the 4 dB pad and stuck it in the box and then ran a ~ 5mm copper wire to connect an SMA output jack, the S22 decreased to -16.3 dB! At 150 MHz — life differs from 7 MHz.

Placing the 4 dB pad between stage 2 and 3 will stabilize + enhance S22, gives a 4 dB higher 1 dB compression point and output intercept, which may make a FBA chain more useful for experiments like transmit driver applications. I can report I measured no oscillations out to 1.5 GHz and this amp works okay for my intended purpose with gain of 26.6 dB at 150 MHz.

I employed 220 pF caps because the plus the self resonant frequency of my 220 pF Murata caps = 154.6 MHz. Therefore, gain at lower HF is low.

Above — FBA breadboard photograph.

QRP — Posdata for October 27, 2013

I first learned radio electronics from SSD, EMRFD and the ARRL Handbook and still remember the joy when my copy of SSD arrived in 1990. Feedback amplifiers litter this book — and for good reason — they offer stable 50 Ω blocks for building up our RF signal. Wes' 144 MHz CW/DSB transmitter still intrigues me and inspired the circuit shown above.

I asked Wes if he kept that 2M transmitter, and if so, could he snap a couple of photos for me? Wes opened it up to see that the VXO and the frequency multiplier chain were missing. Click Click Click . Thanks to Wes, W7ZOI for the photos.

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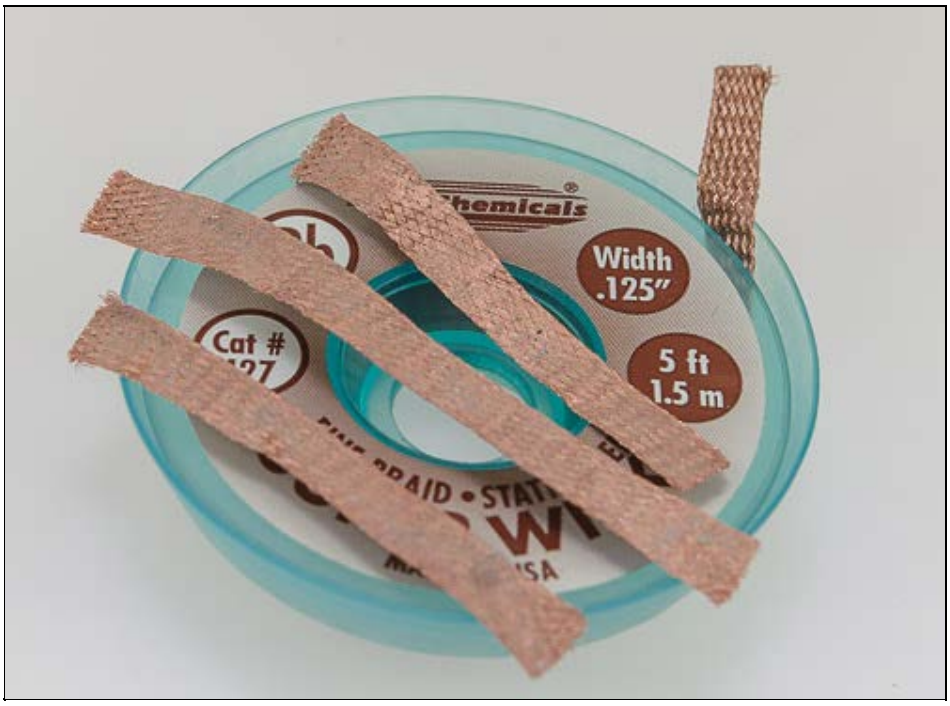
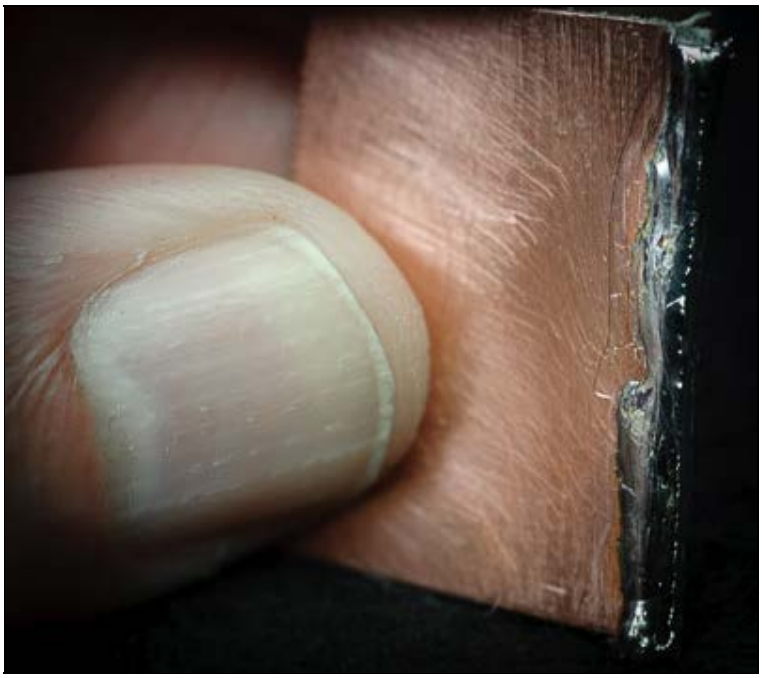
4. Breadboarding Double-Sided Copper Boards

Seeking a low inductance AC ground for some of my VHF and UHF FR-4 prototype breadboards, I join the top and bottom copper surfaces with solder wick along its 4 edges and copper wire vias in the main board area.



Above — A bench staple, solder wick, joins the 4 edges in true Ugy fashion. Many prefer copper foil for this task, but it's not cheap, nor readily available prompting a pragmatic approach. I take a hammer and pound the solder wick [0.125 inch minimum] so it becomes wider and thinner.

Above — Take a piece of wick and tack it in 2-4 places along the top surface of the copper board. The center solder shown above looks perfect — too much solder will flow into the wick and make a difficult bend job around the board edge. Tack just enough wick so you have enough left to go over the edge and to solder on other side of the copper board.

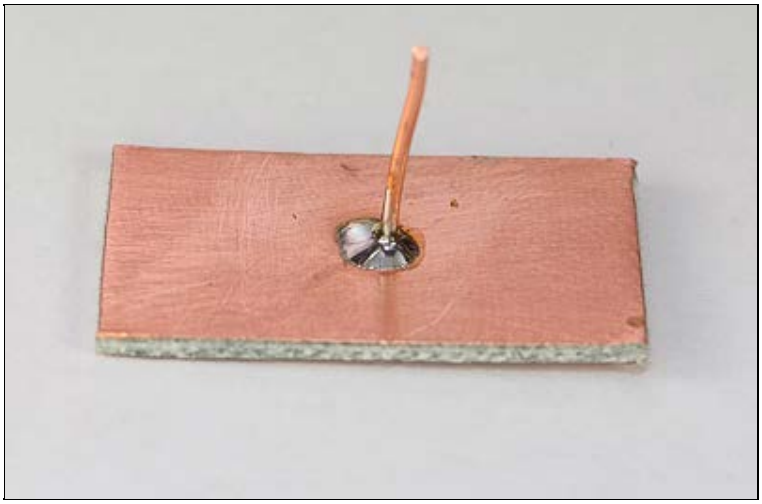


Above — With hand and/or pliers manipulate the wick around the other side to completely bridge the top to bottom. Finally, liberally solder the wick on both surfaces, move to the next board edge and repeat. Click for a rather messy UHF board with copper wick on 2 edges and numerous vias around a prescaler chip, some input + DC lines and a switch.

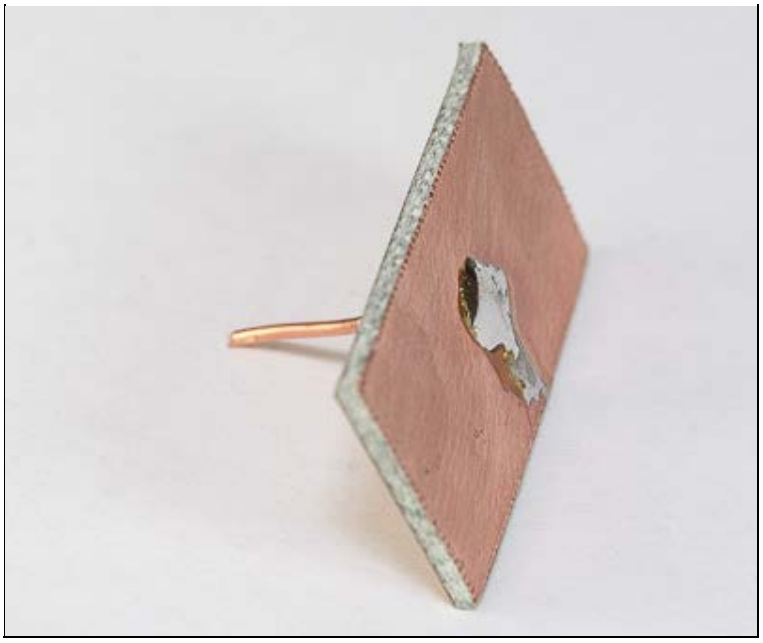
Above HF, most don't make pretty prototype boards to admire — rather, builders strive for good AC grounding and generally stick their board in a metal box for shielding and this hides the breadboard. No one's ever told me I make pretty circuits for any frequency and I'm okay with it.

Onto wire vias:

Above — Your drill chuck will need to accommodate small bits such as those shown above (3/64 inches = 1.2 mm) . I keep these and others that vary from 0.5 mm to 1.2 mm for making via holes.



Above — Magnified via wire soldered at the top side. I found making a hole close to the same size of the via wire will hold the wire in place for rapid soldering. This wire is normally clipped flush at this point. Click for a photo of my flush cutters.



Above — The board shown above was flipped over to solder the bottom copper surface. I try to drill and solder in all vias prior to soldering active devices, since the board heat may damage some parts.

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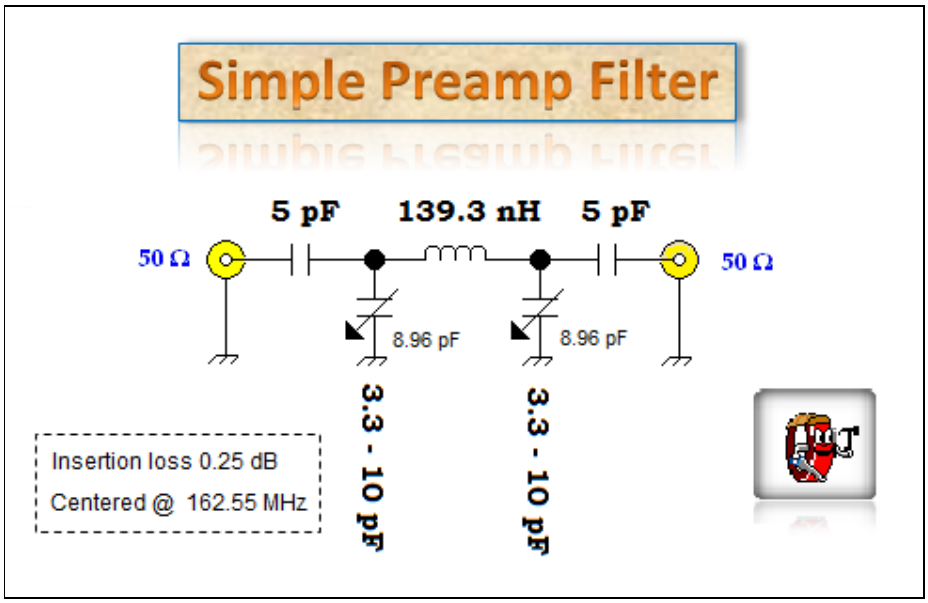
5. Simple Preamp Filter

Some RF filtering should proceed our receiver low noise amplifer (LNA) input — but what? Low-pass, high-pass, band-pass, 2 poles, many poles? Factors informing this decision might include our receive frequencies, QRM, selection of IF and thus our image stripping requirements, plus maximum tolerable signal loss before the LNA. We might not want image rejection from this filter since the post LNA filter can tackle this function.

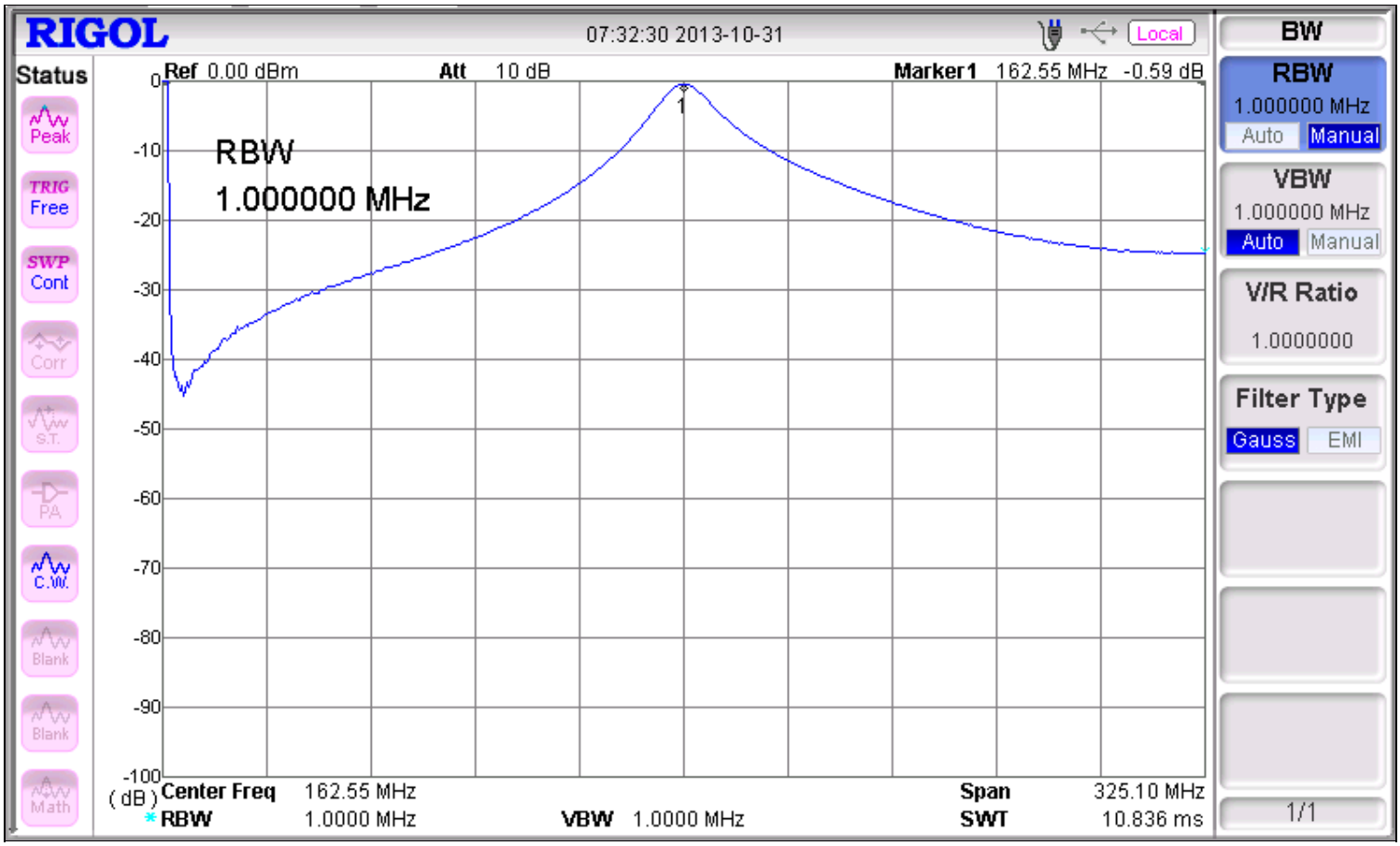
I examined an input filter presented by Joe Reisert, W1JR in the November 1986 Ham Radio. Joe wrote a fantastic column called VHF/UHF World and his context was a high dynamic range 2M receive converter to a 28 MHz IF.

I'm uncertain how to classify this filter: it looks like a standard pi low-pass filter except the input and output series capacitors exert a high-pass skirt. Joe built his filter for 144 MHz. After "building" Joe's filter in Ladbuild08 I tweaked the filter in GPLA *[software from EMRFD]* to center it at our local NOAA weather channel: 162.55 MHz. Click for the GPLA plot.



Above — Input filter schematic with values shown to center this rather symmetrical filter at 162.55 MHz. The half power (3 dB down) bandwidth = 18 MHz. You can easily scale it to other other frequencies — by tweaking the trimmer caps and perhaps squishing or expanding the inductor links some, this filter will tune widely including the 2 Meter Ham band. The IL blew me away (better than calculated) — the resonator Q was high due to the air wound L and Q >= 700 trimmer caps along with good port matching.

Click to view a larger photo of the breadboard. I wound the coil on a 3/8 inch diameter bolt but then spread the turns to get the L and length I wanted. Built on double-sided copper clad board, a few copper wire vias connect top to bottom. I joined the LC circuit to its ports with leaded 5 pF ceramic caps.



Above — Tracking generator plus spectrum analyzer sweep. Under sweep set up, I minutely squished or expanded the L and tweaked the caps to peak the filter at 162.55 MHz while obtaining the lowest insertion loss. Further, I had to tweak the trimmer capacitors to center the filter with the Hammond lid on since putting the lid on changed the center frequency slightly. A *TG equipped* spectrum analyzer proves the ultimate bench tool for VHF and UHF in my humble opinion.

Above — Zero calibration of my AADE L/C IIB meter. By sweeping numerous LC circuits after measuring the L with this tool, I've learned it's reasonably accurate with low inductance coils. I zero the meter prior to *each* measurement and zero calibrate it with the alligator clips butted end-to-end while gripping a small piece of copper wire. Try to keep the same relative alligator clip position during coil measurement. I'm uncertain if this is protocol, but it seems to work for me.

Thanks to Joe, W1JR for this circuit and the opportunity to learn more about component-level VHF design and construction.

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6. VHF Band-pass Filter Experiments

Build Notes

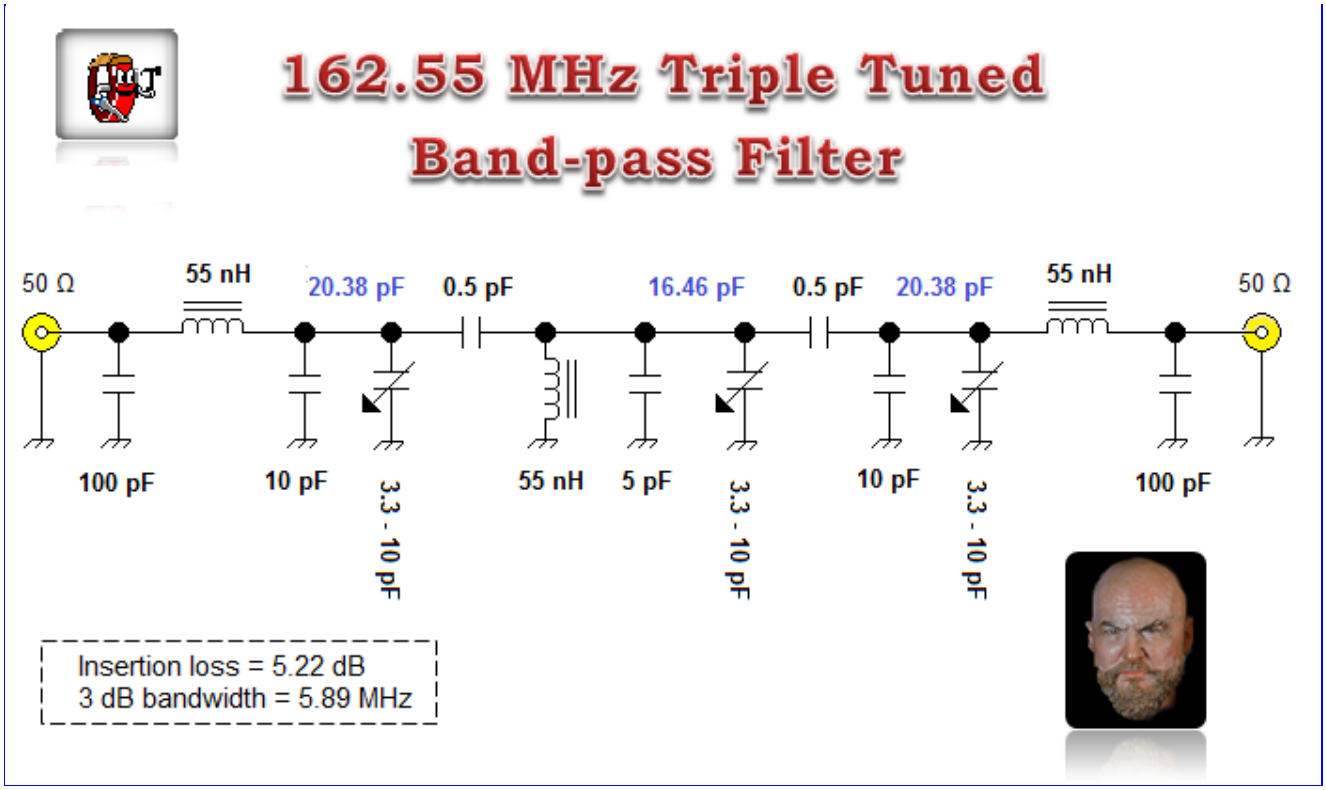
1. All filters designed with the EMRFD ladpac programs DTC08, or TTC08 and then tweaked in GPLA.

2. Seeking a band-pass filter for my experiments with the local NOAA weather channel at 162.55 MHz, I designed, built and tested 3 filters. With a 10.7 MHz IF superhet receiver, the image frequency = 141.15 MHz.

3. My VHF sweep system = a tracking generator plus a spectrum analyzer.

Filter 1

To avoid the need for proper electromagnetic/electrostatic shielding, I kept my HF experimenter hat on and built my first filter with powdered iron toroids (T30-12) knowing I would pay a Qu penalty that may wreck filter insertion loss, bandwidth and port matching. I hoped that the passive electromagnetic shielding offered by toroidal inductors would reduce resonator coupling.

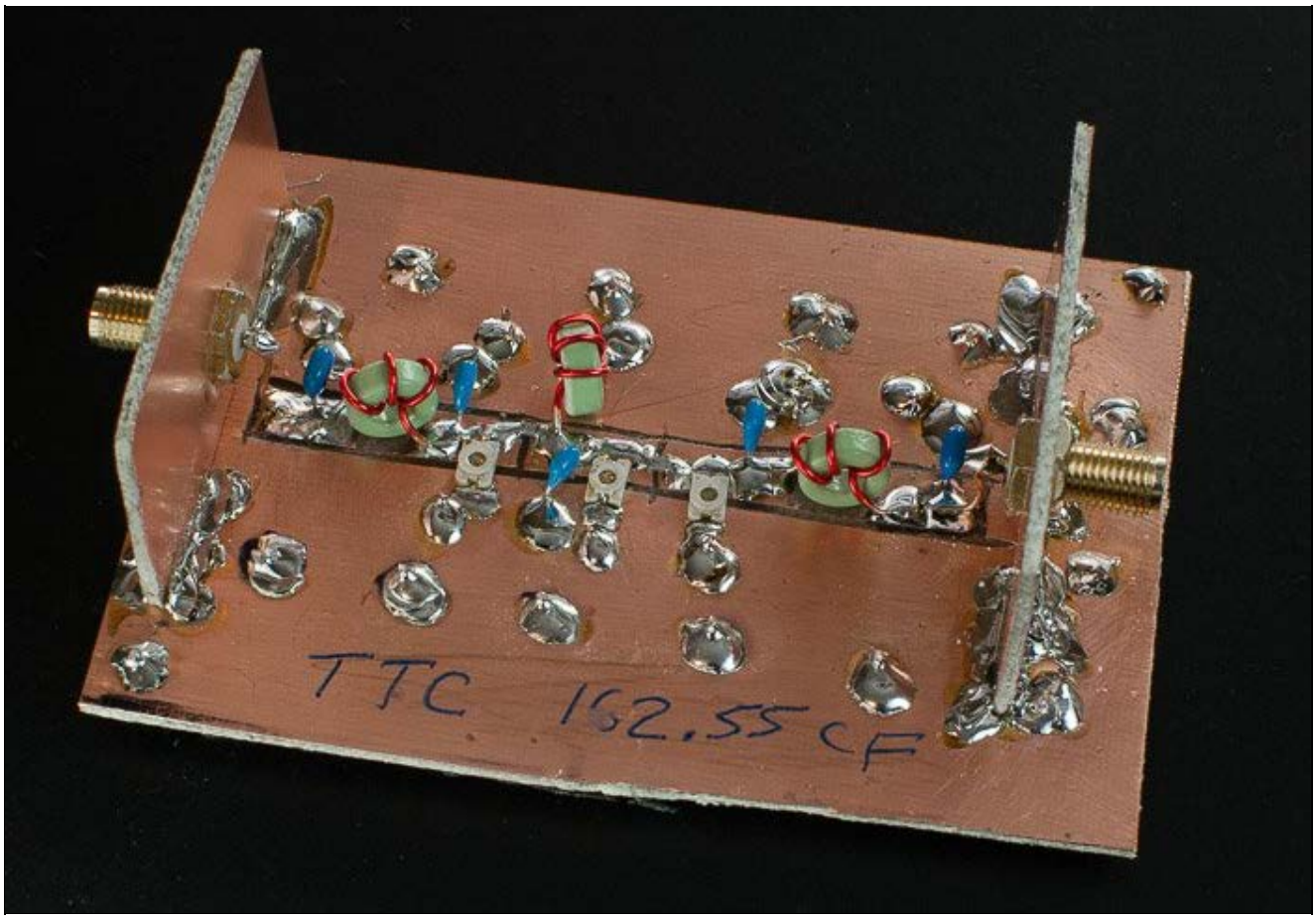


Above — Mixed-mode triple-tuned filter schematic. Click for the GPLA simulation. When we want to go up in frequency plus desire a narrow bandwidth, we face 2 options: reduce our inductance and/or coupling capacitor(s). This taxes our parts collection and breadboard skills since capacitors less than 1 pF are relatively uncommon and require SMD breadboard techniques. I adjusted my design bandwidth and inductance in TTC08.exe to allow the 0.5 pF capacitors plopped in my parts bin to work. For a change of pace, I chose a mixed-mode filter topology.

I strive to place a single high Q variable capacitor in my band-pass filter tanks, however, my Q = 700 SMD trimmers only ranged 3.3 to 10 pF so I stuck a leaded ceramic capactor in parallel with each trimmer.

Common SMD capacitors exhibit low to medium Q and I try to avoid them in my at HF band-pass filters: often I'll just solder a short leaded though-hole caps instead. I also applied this logic to my VHF filter, however, this might prove foolish at VHF since the self-resonant frequency of a given leaded cap is lower than that of an equivalent SMD cap and capacitor XL becomes more significant as we move up in frequency.

Click for a pdf file showing a simple experiment with 3 capacitors . I'll occasionally use ultra-high Q SMD caps in my future experiments based upon these results.

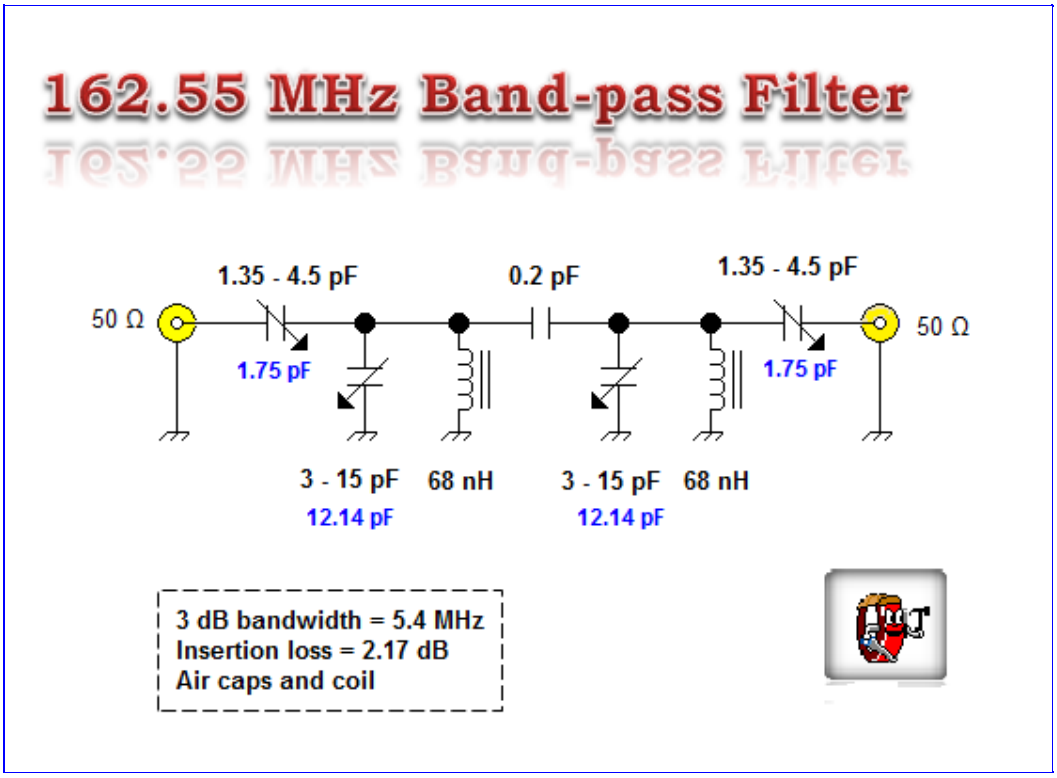


Above — My TTC breadboard. Click for a larger photo. Click for a bench sweep to view a feculent skirt peak in the filter spectogram. Some resonator coupling occurred despite spacing and placing the inductors at right angles. This circuit possessed 6 tweaks: the 3 trimmers and the 3 Ls — I just kept tweaking them until I got the best shape and lowest IL. That's 21 gauge wire on the toroids. My image frequency = 47.28 dB down. Since the sweep yielded such a wide discrepancy from my GPLA simulation, I discarded this filter and moved on to design #2.

Filter 2

Attempting to obtain a resonator Qu of at least 300-400, I moved to air core inductors wound with bare 22 gauge copper wire *plus* air-variable capacitors. Double-tuned classic topology. Without stout shielding, air wound inductors will couple and trash the stopband.

Above — Copper and brass sheets I bought for making filter boxes after Wes recommended using metal sheets or strips for shielding filters. In North America, builders can purchase a sheet metal hand brake plus a rotary cutter/shear for under $100 and make all sorts of boxes for radio gear. I'll equip my QRP workshop with such equipment over the next year. This was my first experience man-handling sheet metal or strips.

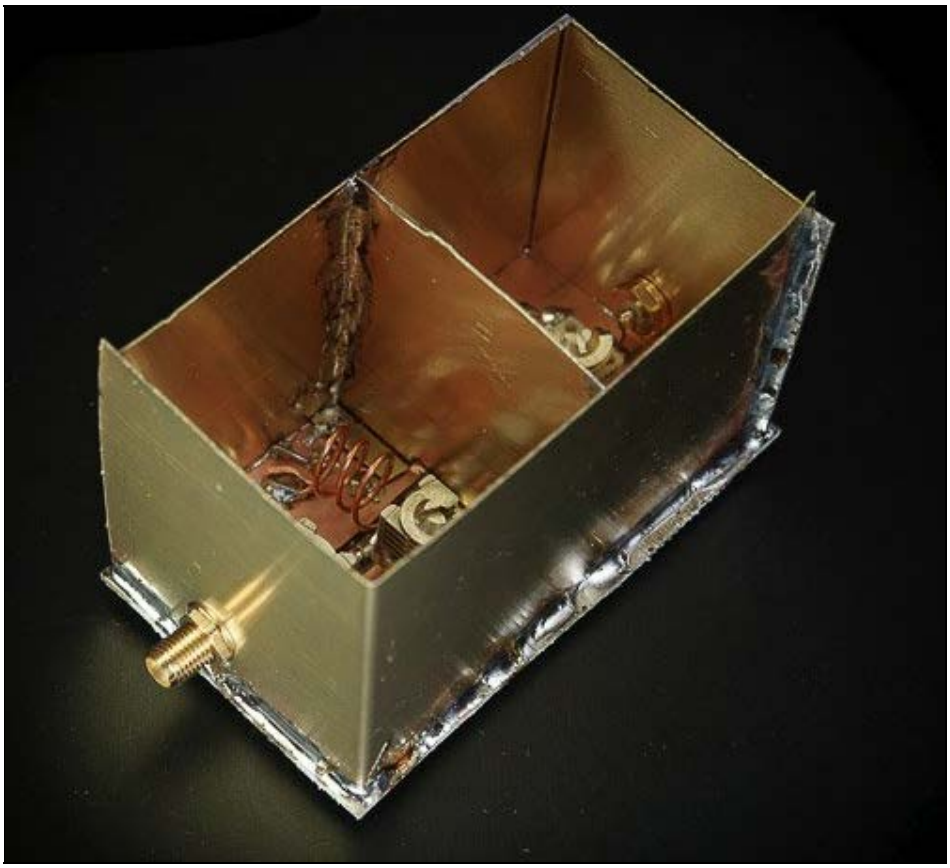


Above — Schematic diagram of Filter #2 with tabled IL and half-power bandwidth. Click to view a macro photograph of the filter. I soldered the brass box onto a double sided copper-clad FR4 bread board laden with some copper wire vias to enhance the ground plane. The 0.2 µF coupling cap = an ATC microwave cap.

Click and click for the sweeps. Notice the splendid skirt symmetry at the filter top and also the fantastic peak attenuation in the low-pass skirt

(~81 dB from the filter peak to the lowest point in the low-pass filter skirt). I'd never seen deeper attenuation on my bench before: quite gratifying. My double-tuned filter had essentially matched the 3 dB BW of triple-tuned filter #1 with less insertion loss.

Not all appeared perfect however; attenuation of frequencies above the passband looked so-so and I was unable to get a lower 3 dB bandwidth by manipulating L and C in each tank.



Above — My first ever brass box soldered on a 2-sided copper clad board: Filter #2. Lacking proper tools and also the knack for making beautiful chassis like Dave, AA7EE does, I just did my best. My only concern was getting tank isolation. I chiseled out a small cut in the center brass divider just high enough to clear the SMD 0.2 pF chip capacitor and center copper strip. Overlaps allowed soldering of brass walls After tweaking the L and C parts, I placed a brass lid on top during sweeps. Okay, onto filter # 3:

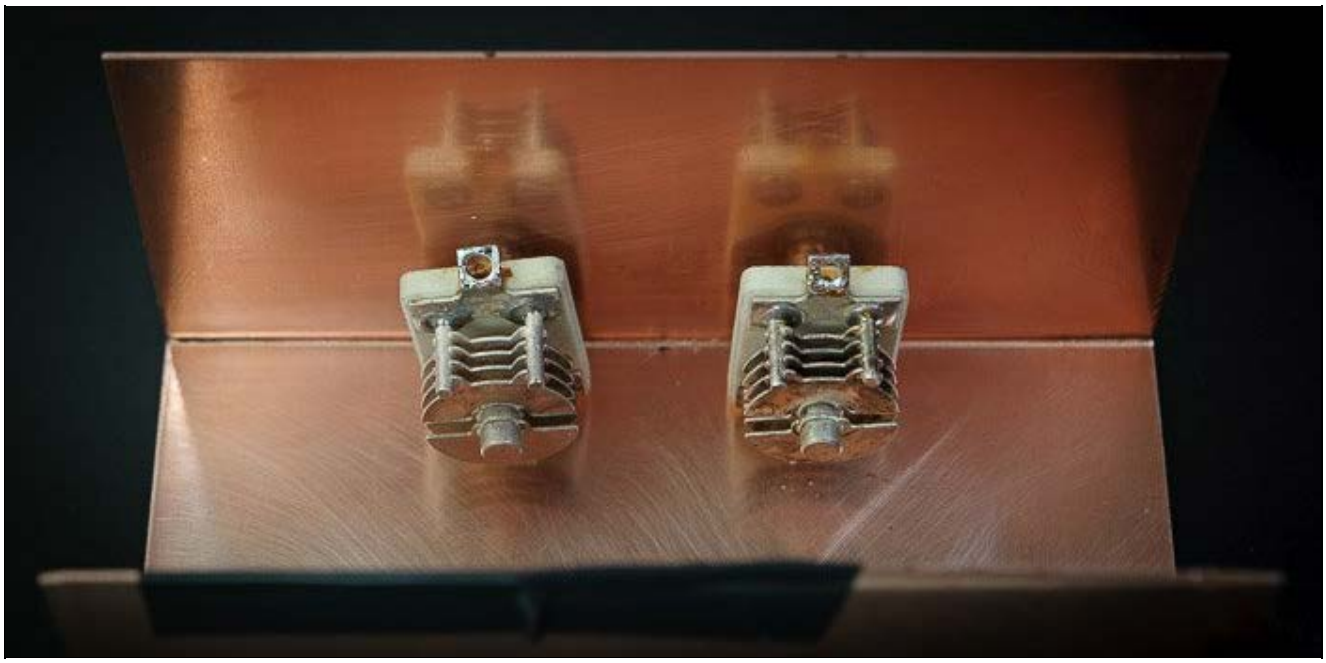
Filter 3

When coupling caps fall below 1 pF, some builders place their resonators in close proximity to couple the tanks, however, these usually involves slug-tuned powdered iron or ferrite inductors with a generally low Q. The coil consists of cup core and a threaded center core which combine to give a magnetic shield. A tin plated copper can surrounds the core for electrostatic shielding. I think after these experiments — placing each resonator in its own sealed box seems a better option.

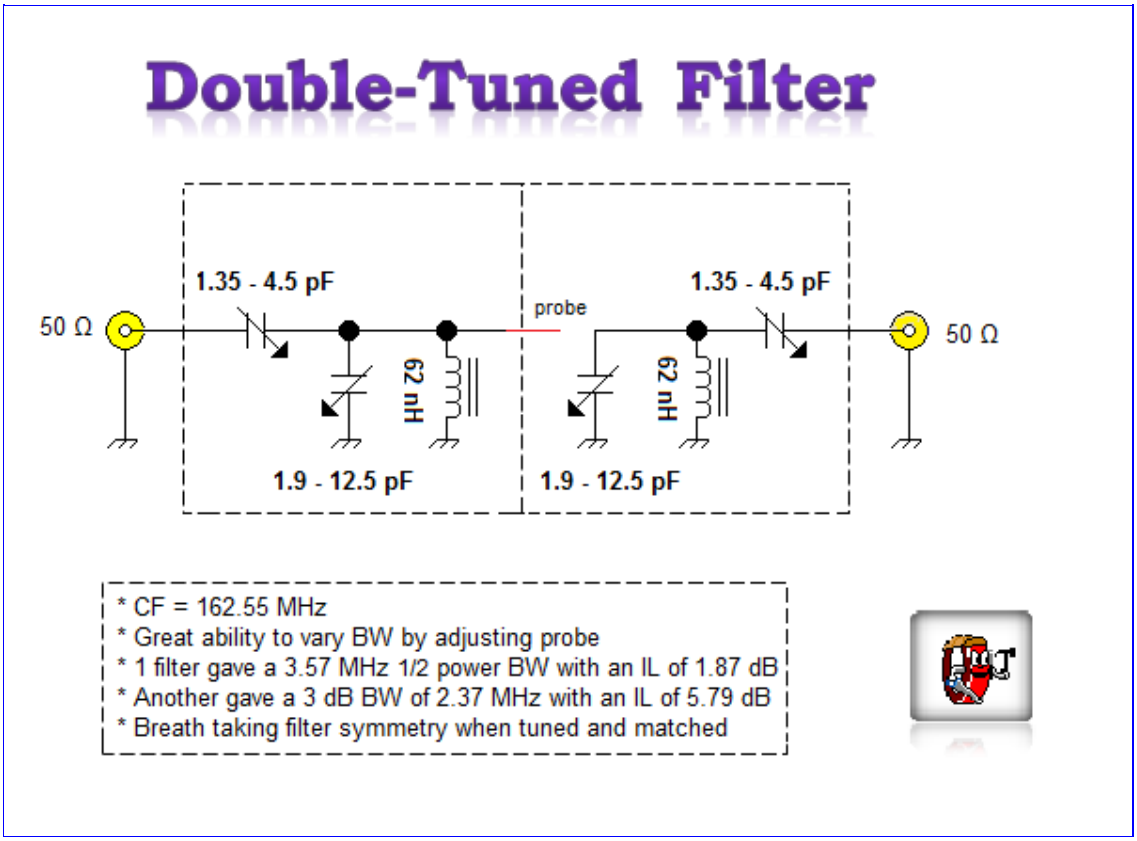
I noticed another approach in The Double Tuned Circuit: An Experimenter's Tutorial in QST for December 1991 by Wes, W7ZOI. Wes drilled a hole in the tank compartment divider and passed a wire soldered to 1 tank through the hole and positioned it near the variable capacitor of the other resonator. We change coupling by adjusting this wire.

The wire may be thought of a "tweakable" capacitor, however, the key point is we need to couple energy from 1 resonator to the next; whether by wire, capacitor or inductor, proper resonator coupling will give the required skirt shape within the the limits imposed by Q and the matching to the 50 Ω ports. When establishing a filter passband, the end Q and resonator coupling pretty much dictate the outcome.

Filter #3 involved more mechanics than electronics. I made a U shaped box from a small sheet of copper.



Above — My copper box drilled for the 2 air-variable tuning capacitors. I built the walls, top and divider from thinner gauge brass strips. Rosin flux aids soldering your metal sheets together — I soldered mine with a Weller 80 Watt iron and began with the center divider. Click for a side shot after completion.



Above — My filter schematic with some tabled data. Click for a top off photograph where you can easily see the black wire probe. To change coupling, you change the wire length and/or its distance from the second tank capacitor. I really enjoyed moving the wire and watching the outcome on the screen. Per usual, I tweaked the coupling, Ls and end caps to derive many 3 dB bandwidths. In 1 arrangement with a 3 dB bandwidth of 3.57 MHz, the insertion loss was only 1.87 dB — I had evidently found a perfect ratio + combination of L and C yielding a low IL.

I'll increase the coil wire gauge in future experiments to lower solenoid resistance and try to garner even more Qu.

This is the *best* filter I've ever built. After peaking the tanks to 162.55 MHz, resonator adjustments weren't really needed. By moving the wire probe I saw that BW and IL change inversely. While tweaking the series caps and moving the probe, it felt like déjà vu from my experiences simulating these exact changes in GPLA over many years — except now they were alive and kicking. With some inductor manipulation (or as

needed, changing the inductors) this filter can be centered from ~140 to well over 200 MHz. A high-octane, dramatic and versatile filter indeed.

Click for a sweep where I moved the wire very close to the neighboring tank capactor and over coupled the filter.

Click for a screen shot of another sweep.

Conclusion

With new awareness that resonator coupling and other issues may *also* occur in HF band-pass filters, making these 3 filters surely boosted my future band-pass filter construction standards at any frequency. Over the years, builders have emailed me sweeps of HF band-pass filters (usually VNA sweeps) with also sorts of extra resonances +/- poor insertion loss. Since I'm just an amateur, often nimrod RF enthusiast, I won't pretend I possess all the knowledge to critique their filter woes.

However, photos of their HF filters usually look messy and rushed: opposite to our needs — eschewing important factors like symmetrical layouts, inductor spacing, a first-rate ground plane, larger gauge + well spaced wire wound on big as possible toroid cores Yikes, I better stop now lest I venture into filter folklore.

Today, designing a filter with software is merely half the battle — construction poses the harder task. Unsurprisingly, measurement and experimental methods will elevate your outcomes.

Filter analysis with a sweep system is the ultimate, however I built and then tested many band-pass filters with a frequency counter, a signal generator, a 'scope and a return loss bridge during the first 12 years of my RF life.

Assuming you have a filter that peaks and isn't overcoupled: to measure the 3 dB bandwidth, first peak the filter to resonance for the highest peak-peak voltage in your 'scope. Here's the rest of the procedure:

1. Multiply the peak-peak voltage of your centered filter by 0.707 and record this value.
2. Decrease the frequency of your RF generator until the scope shows the .707 value and record it as Frequency Low
3. Increase the RF generator frequency above the peak until you once again get 0.707 of your peak-peak measured voltage. Record this as Frequency High.
4. 3 dB bandwidth = Frequency High - Frequency Low.

You may also wish to measure the insertion and return loss as described in EMRFD, Radio Society Handbooks, or the RF Workbench series on this web site.

Whether you run superhet, low IF, SDR, or Zero IF (DC) receivers, or need to filter a transmit chain, well made band-pass filters just might boost your project performance. VHF poses additional challenges but offers great learning opportunities.

QRP — Posdata for Nov 12, 2013

A. Notes for Filter #2

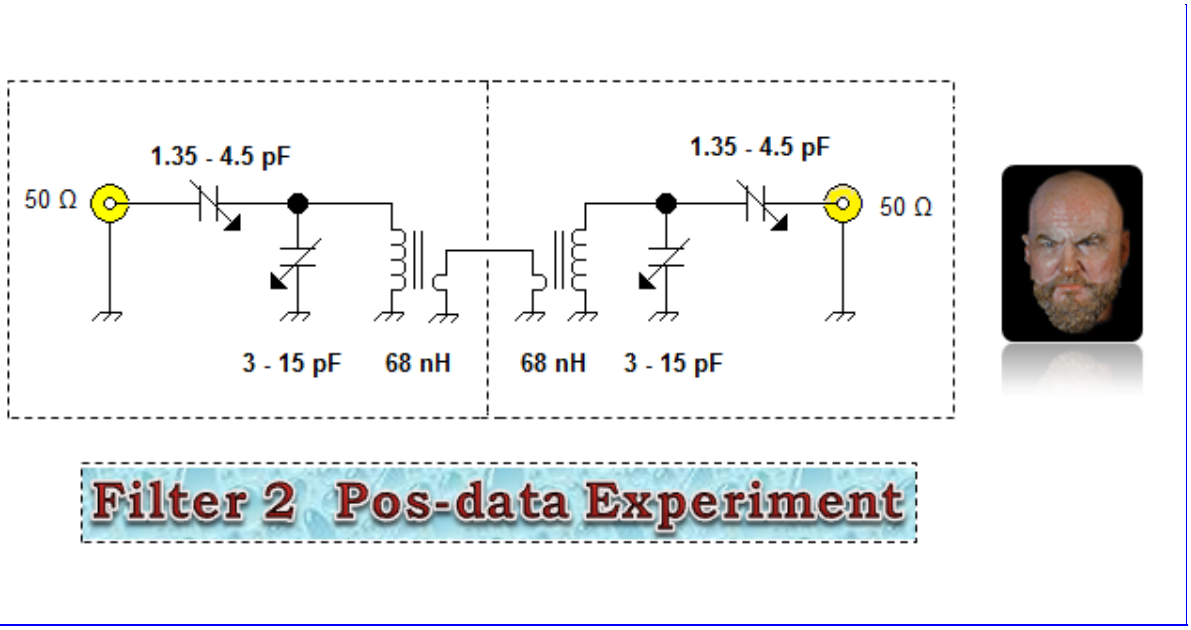
Victor, 4Z4ME wrote to explain why the attenuation above the passband in Filter #2 looked mediocre. Filter topology and component values yielded the poor high-pass attenuation and not unwanted coupling as I originally shared.

Victor wrote: "The problem is that all the couplings are of one nature, in this case capacitive.
At very low frequencies the inductors short the signal to ground while the capacitors impedance gets higher so it is clear that the output signal goes to zero.

At high frequencies the inductors may be regarded as disconnected because their impedance get very high, however although the capacitors impedance get lower , you have them both at the series arms and at the parallel arms, so the signal gets to be attenuated by their impedance ratio and not get shorted to ground.

This can be shown in analytic calculation, or easier by simulation. See the attached file (LTSpice simulation). You can see that the filter (similar to your filter) gets closer to a finite attenuation at high frequencies, getting closer to the attenuation of a similar circuit with the inductors *removed*.

You can get better high frequency results by trading the center 0.2pF capacitor coupling to an inductive coupling. Remove the coupling capacitor and insert a wire through the center shield, ground its ends close to the main inductor. Shape the wire to a small loop near the grounded end of each of the resonator inductors to get small inductive coupling, and you get again a good BPF but with better high frequency attenuation. Instead of loops you can connect the coupling wire at a tap at the bottom of the main resonator inductors, close to the ground".



Above — Following Victor's suggestion, I inductively coupled the Filter #2 resonators with a wire looped near the grounded end of each tank inductor. Please view the schematic above. High-pass attenuation above the passband improved significantly. Click for a sweep out to ~ 511 MHz that shows the peak high-pass attenuation much closer to the low-pass skirt peak reponse.

At 2X center frequency (325.1 MHz) the high-pass skirt peak attenuation closed within 14 dB of the low-pass peak attenuation compared to the original Filter #2 response where the delta was around 30 dB. Further improvement occurs as we move above ~400 MHz. Big thanks to Victor for his support.

B. Sheet Metal Safety

A cut finger while working sheet metal prompted the following remarks: Jagged edges from cut sheet metal (especially brass stock it seems) may lead to deep cuts. Filing a burred edge with the metal clamped in a vice between some thicker metal stock with its raspy edge barely showing will reduce the "knife edge effect". When drilling, clamp the metal piece to your bench on a wooden backing board . If you jam your hand drill in an unclamped piece of metal it could spin and really gore you. Safety first.

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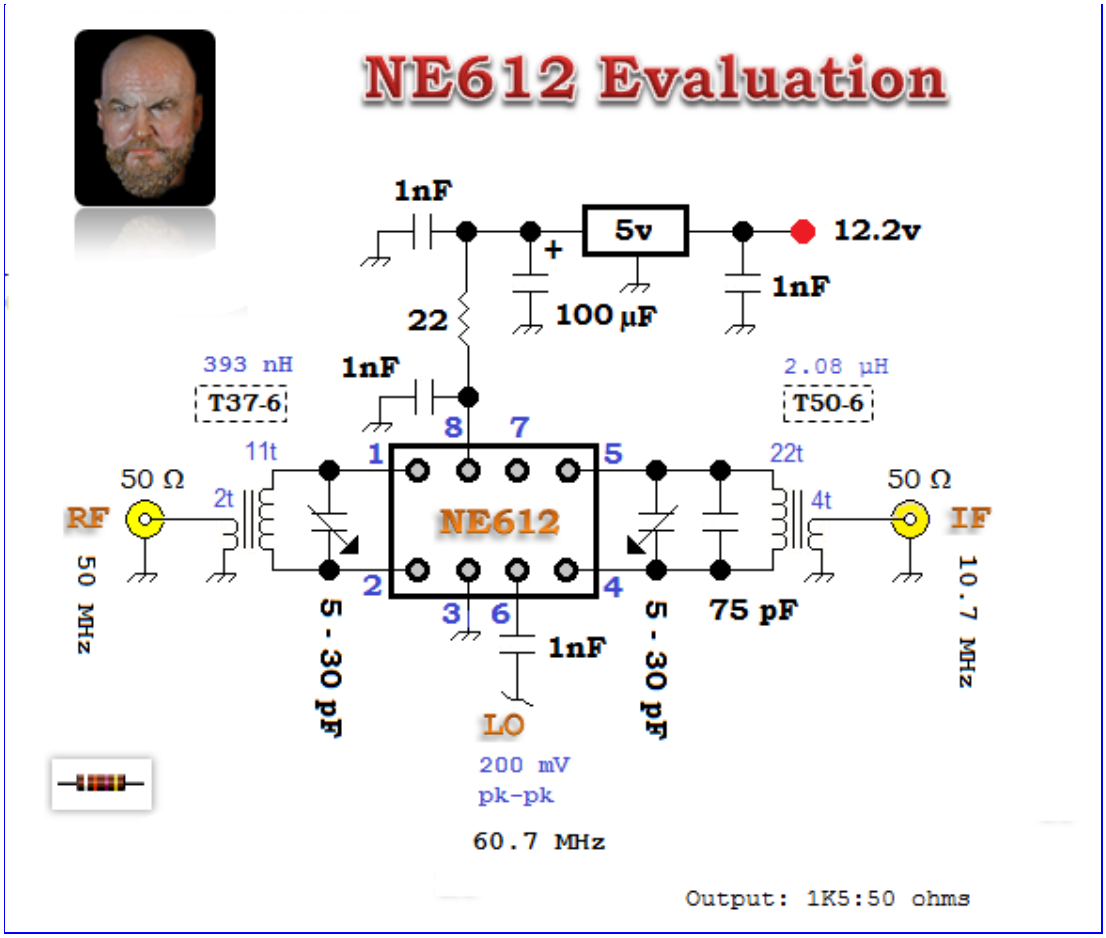
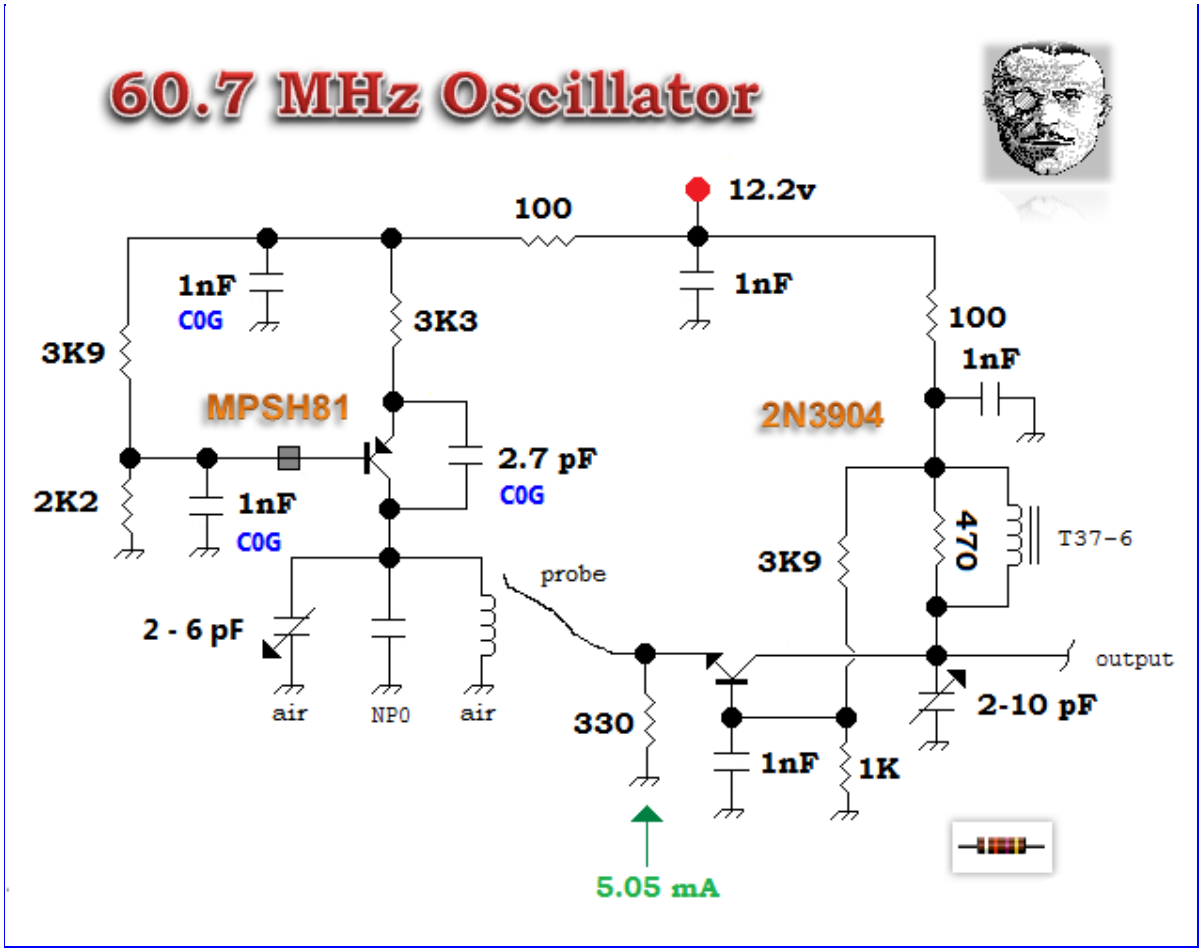
7. NE612 Mixer Diddy

The Signetics NE612 and ilk enjoys wide use by the amateur radio community in novelty-grade Ham band receivers, or in somewhat better receivers laden with abounding front end filters + *stout,* switchable RF attenuators. Do you think of the NE612 as an ersatz receiver mixer? Some only consider them for the transmit mixer slot. Read this link why.



Above — My base schematic. I applied differential input to cancel 2nd order mixer products and balanced output to obtain maximum conversion gain. Click for the breadboard. Powdered iron toroids, Q = 1500 trimmers and a fixed C on the output formed the resonators. I chose a 10.7 MHz IF to drive low-cost ceramic filters in further (unshown) experiments. Normally, we don't bother tuning the input since our bandpass filter should launder the input signal sufficiently.

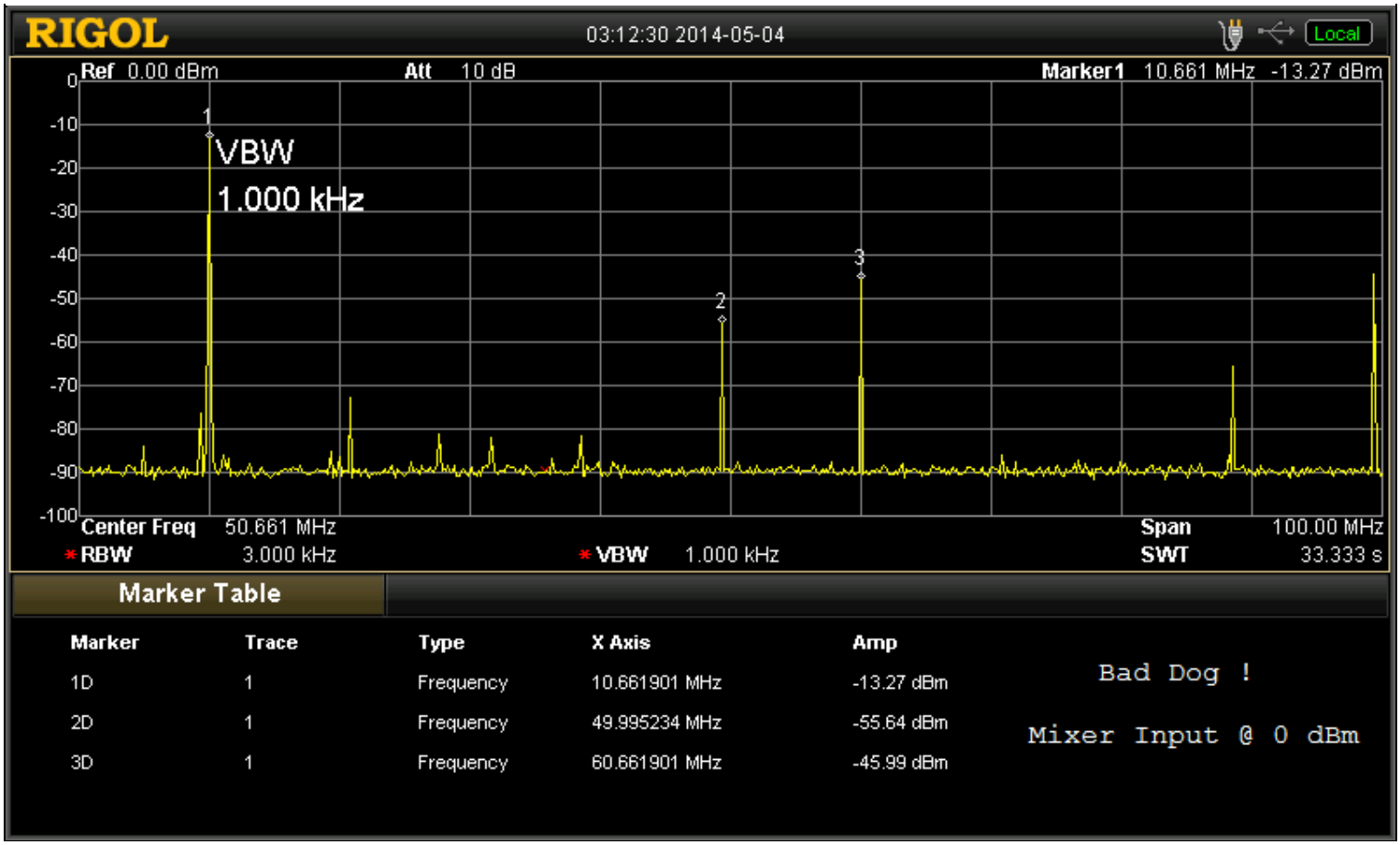


Above — My 'thrown together' 60.7 MHz local oscillator. A wire on the common base buffer amp input avoids the series R and C usually required and I moved the wire probe to set the output power between 180 and 250 mV peak-peak into a 1K resistor load.

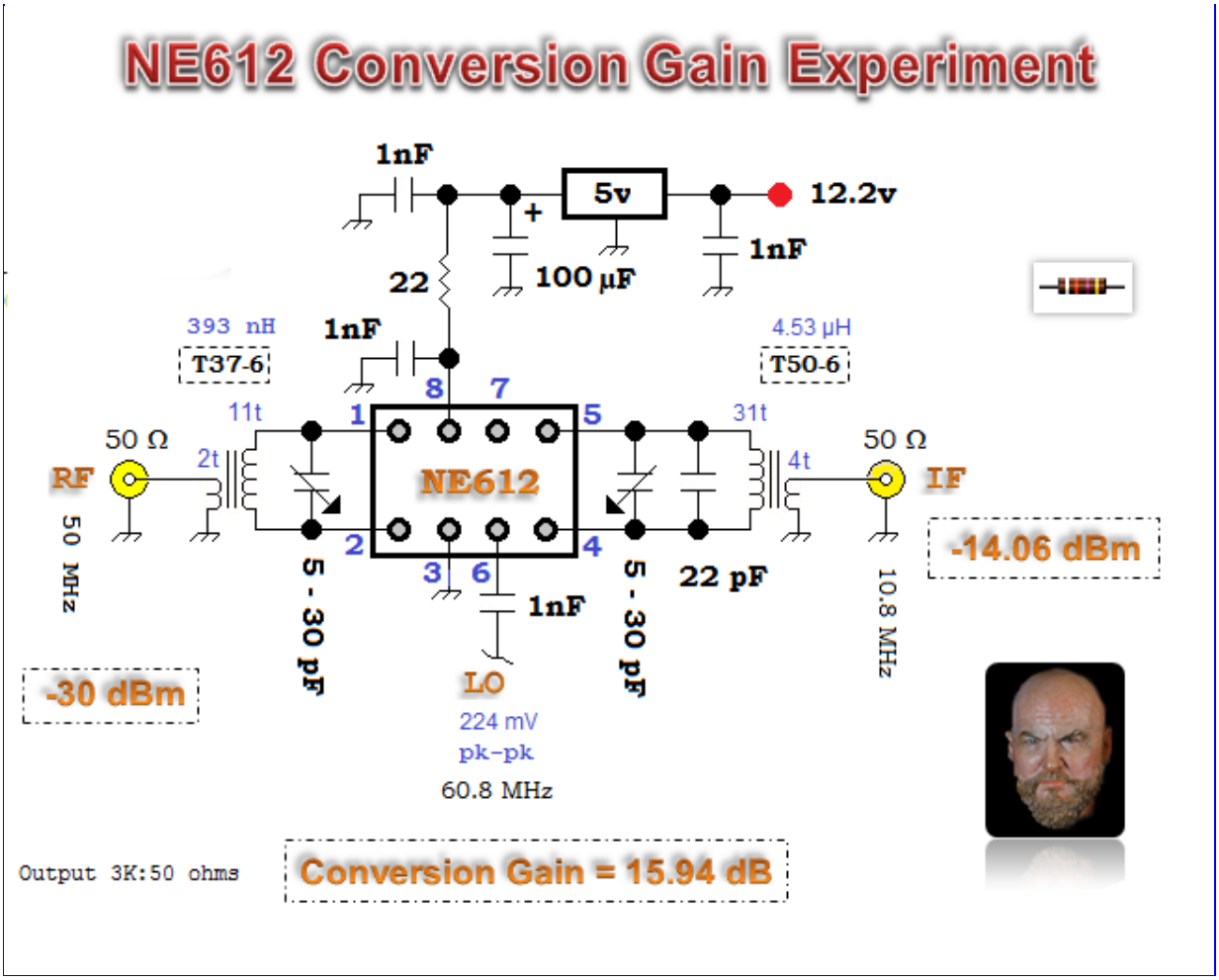
Input power	IF	RF	LO	2RF	Gain
-35 dBm	-24.3 dBm	-68.71 dBm	-47.09 dBm	-68.14 dBm	10.7 dB
-30 dBm	-19.89 dBm	-63.68 dBm	-47.01 dBm	-58.29 dBm	10.11 dB
-25 dBm	-17.41 dBm	-60.07 dBm	-46.96 dBm	-50.06 dBm	7.59 dB
-20 dBm	-16.21 dBm	-55.62 dBm	-46.65 dBm	-45.06 dBm	3.79 dB
0 dBm	-13.27 dBm	-55.64 dBm	-45.99 dBm	-53.28 dBm	-13.27 dB

Above — A table showing the mixer input, IF, RF, LO, 2xRF [the worst spur] power readings, plus calculated conversion gain. This table echos the datasheet + work by others — input power >= -25 dBm drives this mixer into compression. The NE612 runs low current plus low voltage across the output transistors — thus the maximum output power is limited to somewhere between +2 dBm to +5 dBm from my experiments.

Even with -35 dBm input power, I measured a conversion gain much lower than the typical 14-18 dB enjoyed by others who applied differential output at ~ 50 MHz. Turns out, I goofed — each output transistor contains a 1K5 Ω resistor, so when wired differentially, the output Z = 3K. Back to the bench!



Above — SA of the mixer output with 0 dBm RF input. Something we would never do in real-life, but I wanted to see the spurs with such a drive: the spurious output actually looks great! The NE602, NE612, MC1496 and SO42P [all Gilbert cell mixers] output cleanly; especially when compared to diode-based mixers.



Above — My mixer with output Z corrected at 3K Ω. Wow, a conversion gain of 15.94 dB with -30 dBm drive. All good now. Click for the IF measurement on my SA. The LO had drifted up to 60.8 in this experiment, however, I left it and peaked the output resonator for maximum smoke. I also extracted another 0.5 to 0.8 dB gain by tweaking the LO drive. In my breadboard, a drive of 224 mV pk-pk gave the highest conversion gain.

Hereafter, I'll drive my NE612 transmit mixers with between -26 to -30 dBm on the RF port with differential input and output.

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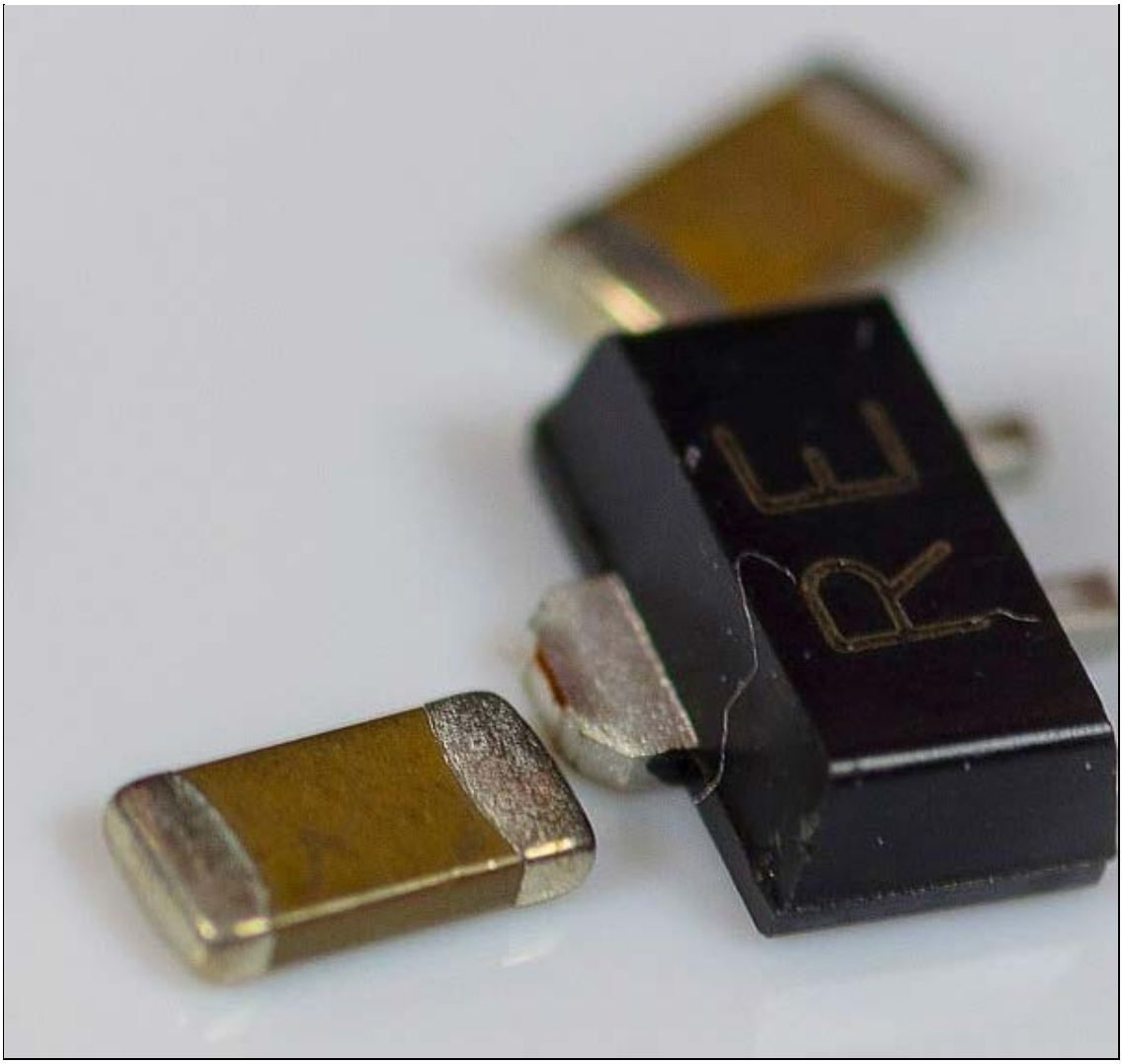
8. Miscellaneous Photographs



Above — Hair or lint in my SMT macro photo

Above — Instrument grade N-male to SMA female

Above — Instrument grade N-male to SMA female; gorgeous machine work.

RF Workbench Page 6 — QRP Transmitters

RF Workbench 6 explores some questions regarding basic transmitter power amplifer (PA) design. Since I write for both Ham and non-Ham readers, I feel badly that RF Workbench 6 excludes pure radio listeners; please consider getting your amateur radio licence to join the fun.

For Class A, AB and C transmitters in current source operation, radio builders knew most everything presented on this web page back in the 1980's — humbled and guileless, I've but a skosh of new ideas to further elevate this topic. Sometimes I feel bound in a creative straightjacket because puzzling design questions paralyze my understanding and progress. This web page explores a few transmitter questions born on my QRP work bench and I hope you'll relate and go design your own high-octane transmitters. As always, I hope to share insightful + actionable info for the DIY scratch builder; albeit from a lay-person perspective.

I borrow heavily from the work of Wes, W7ZOI per correspondence, direct contributions, from EMRFD and the references listed in Section 5. My sincere thanks to Wes for his support and inspiration.

Preface

You require an amateur radio operator licence to operate a radio transmitter and it's up to you to know and ensure your transmitter spurious emissions meet or beat the regulations in your country.

Although I measure my circuits with calibrated equipment, expect your measured voltage and current to differ from mine. **Why?** Variables may include your DC power supply voltage and current rating, dummy load resistance, stray L and C, cables and wire effects, bench errors, parasitic oscillations, temperature effects and parts tolerances.

I unabashedly declare my love for experiential learning: to build and measure, discover and develop — and best of all, to tune, tweak, and transform my RF power amps to unleash the "smoke" and power we all crave.

Contents

Section 1: 21.06 MHz Bipolar Transistor Transmitter in Class C: Dig out your old transistors and explore some Class C topics.

Section 2: PA Measurement with Examples: Measure collector V and I

Sections 3, 4: Incomplete and coming in 2013-14. I'll post any additions on the QRP Log -- Thanks

Section 5: I'll list all the references for this web page

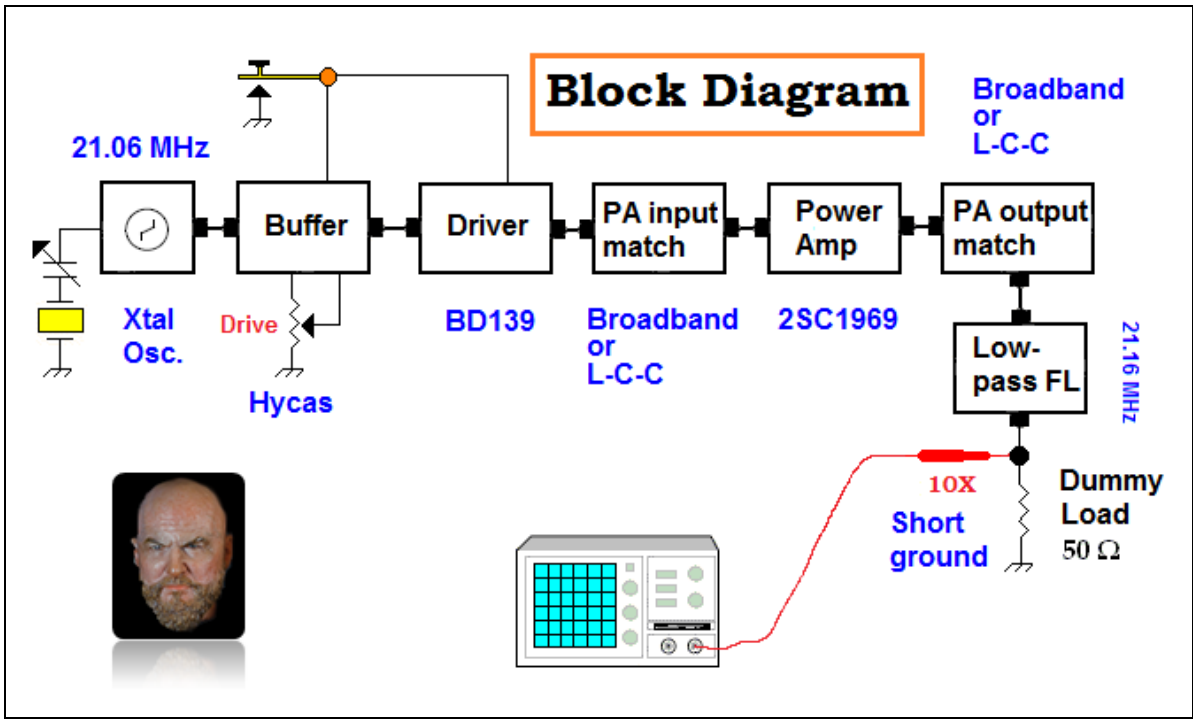
Section 6: Photographs and Miscellany. I'm told at least ~~3~~ (now 4) people enjoy my photos.

Section 1: 21.06 MHz Bipolar Transistor Transmitter in Class C

Despite all the buzz about PA efficiency and the trend towards more efficient switched-mode power amplifiers, my email inbox fills with content regarding those Class A to C power amps. Consider too, most Hams, including QRP buffs, power their rigs with an AC connected DC supply that makes battery life moot. Perhaps our love of PAs in current source operation harkens from our historic tube days: do we simply enjoy making heat and wasting power? Like me, you might just love making and measuring bipolar transistor circuits.

A Class C power amplifer excited by a crystal oscillator offers both new and reawakening homebrewers the chance to make a simple but relatively powerful, 1 frequency transmitter with only 2 - 4 RF stages. Keying a Class-C PA embodies the Ham glory days: put some RF into the

sky with basic, analog circuitry.



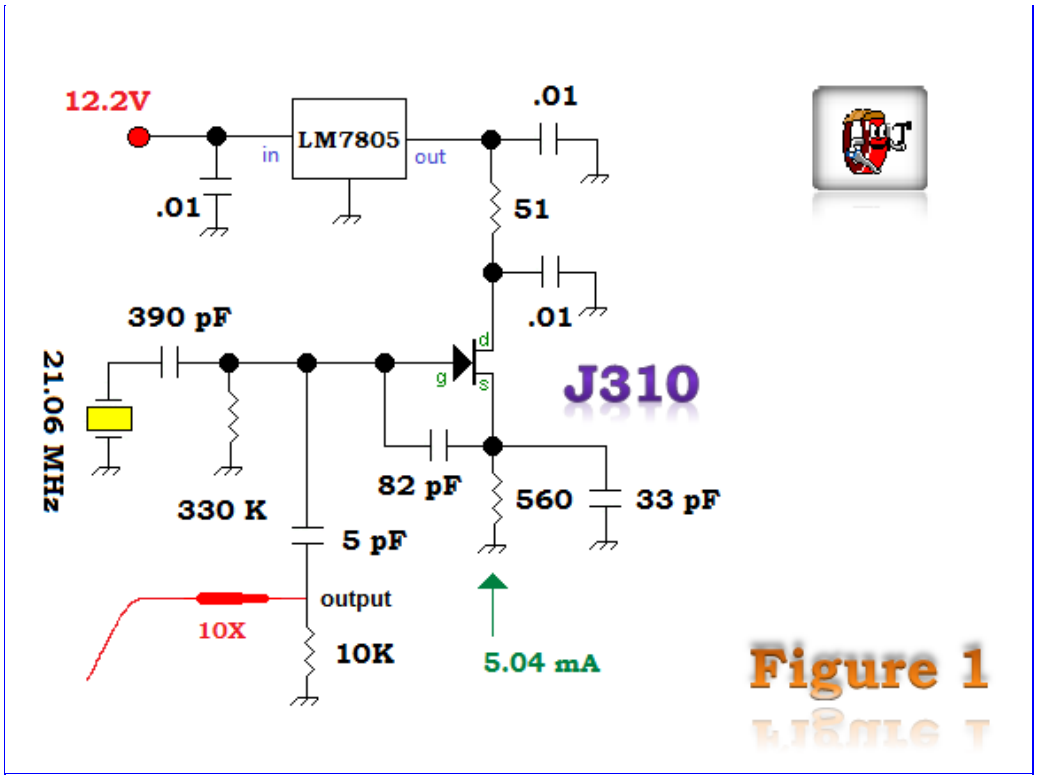
Above — The block diagram for all the experiments in Section 1. Each stage involving transistors was built and tested from left to right before making the next stage.

Crystal Oscillator, Keying Switch and Hycas Buffer Amplifer

As more a listener than a talker, I wanted a low-level crystal oscillator to keep my *key up* current draw low. In the class C power amplifers based on the above block diagram, I measured only 7.5 mA. *key up* and several hundred mA *key down*. I also wanted my back wave to be > 40 dB when I eventually converted the PA to Class AB during another set of experiments.

Design Question 1: What backwave should I strive for?

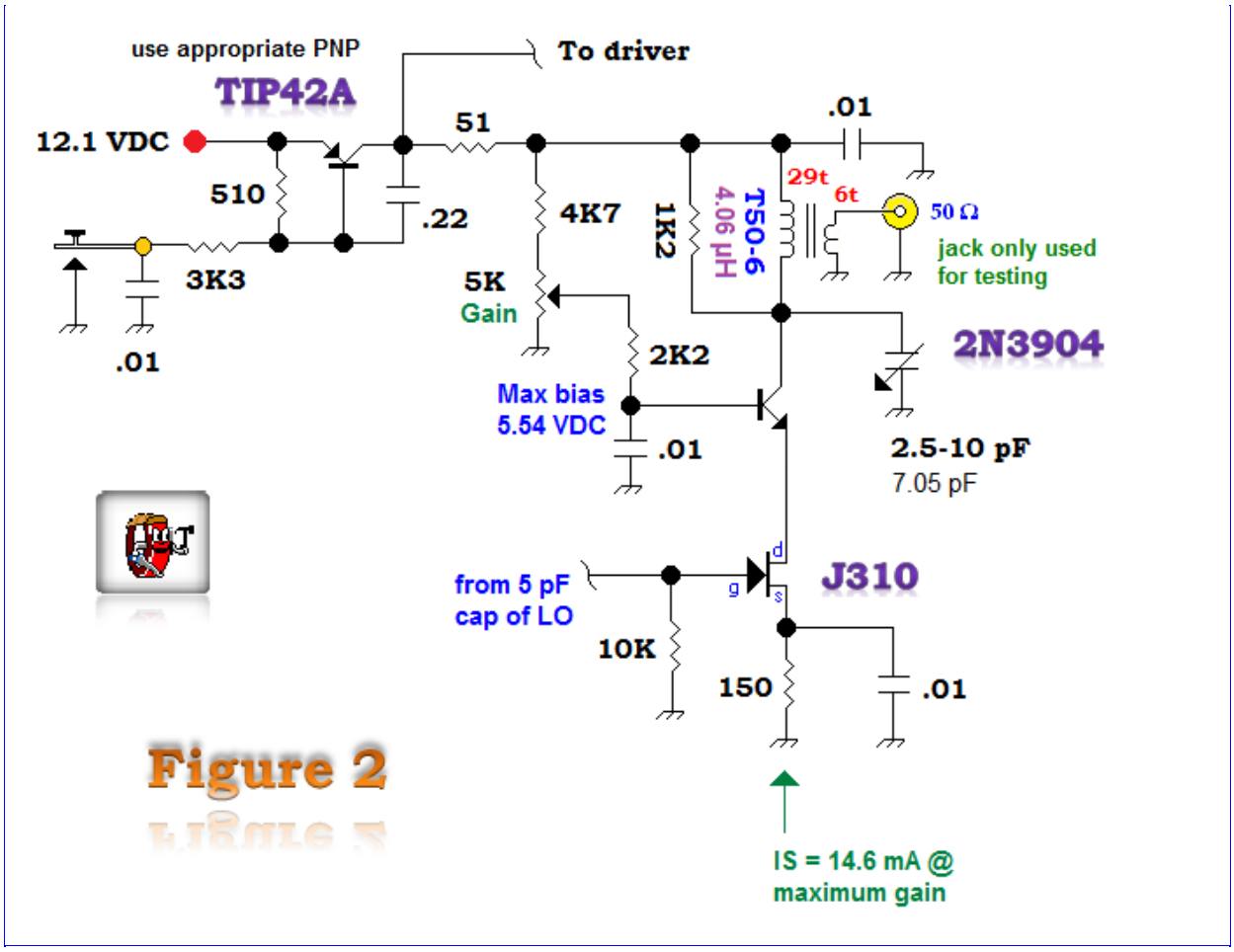
Backwave = the dB difference in output power between key up and down. Opinions differ and context may partly explain why. If you're working DX, backwave is less important than when your working nearby ops on 80 meters. From asking respected authors, experiments and operating, I feel 40 dB should be the minimal backwave to strive for. Operating Class C, your final(s) are turned off during key up and a >= 40 dB backwave proves an easy goal.



Above left — The 21.06 MHz crystal oscillator. Click for the output voltage into 10K Ω. Click for the FFT. Seeking low distortion and output power, I placed variable capacitors in the feedback and source slots and also played with the 560 Ω source resistor. While watching in my 'scope and tweaking , I chose the the lowest distortion signal under 1 Vpp. The variable capacitors were removed, measured and replaced with nearest standard 5% parts. The 10K load of Figure 1 corresponds to the gate resistor of the hycas amplifer input.

Above — Without the 390 pF capacitor, the crystal oscillator vibrated at 21.058 MHz.

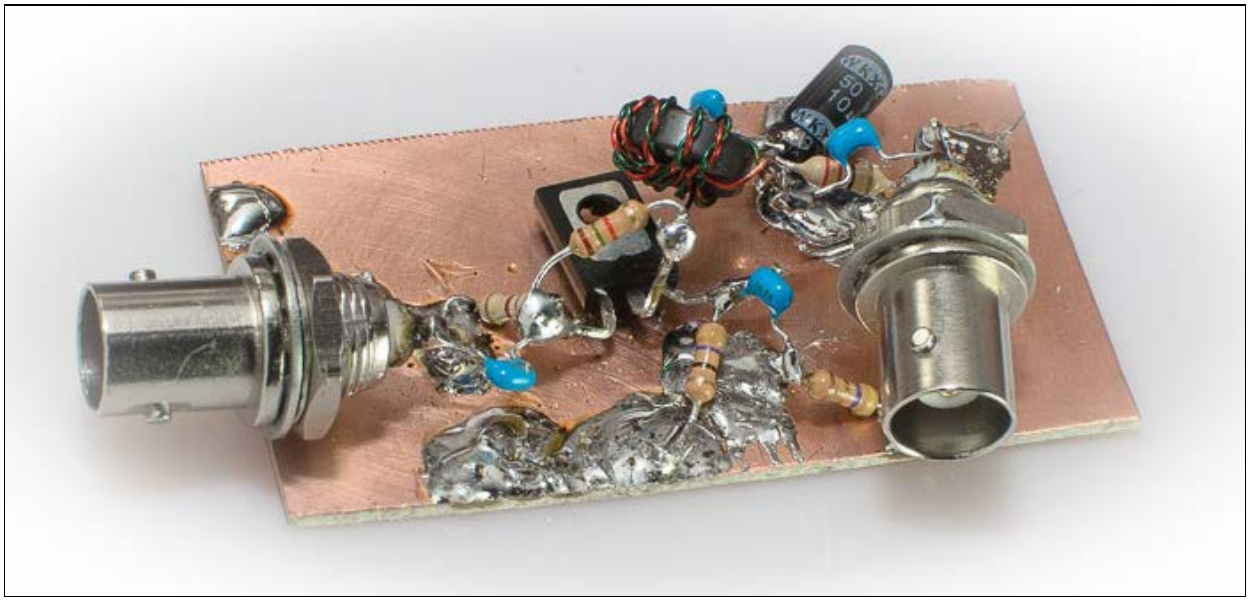
Above — The crystal oscillator measured frequency after experimentally adding a 390 pF cap. Should I have tried a 470 pF capacitor and attemped to get 21.0600 MHz?
My oscillator looks unconventional and in some cases you might want 1 with higher output voltage. Check EMRFD and the other W7ZOI references in Section 5 for xtal oscillator design ideas.



Above — Figure 2, the hybrid cascode buffer. I've never tuned, nor keyed a hybrid cascode amp before and it works well. Click for the oscilloscope output. Click for the FFT measured with a long wire grounded 10X probe. I applied a junkbox TIP42A switch lying on my bench since I keyed a number of *unshown* high current stages during my experiments. Only about 10% of my experiments ever make it onto the QRP/SWL HB web site. For the PNP switch, just use whatever can handle the *measured* current with key down: 2N3904, 2N4403, 2N2907, BD140, etc.,etc..

Yes dear reader, measure everything.

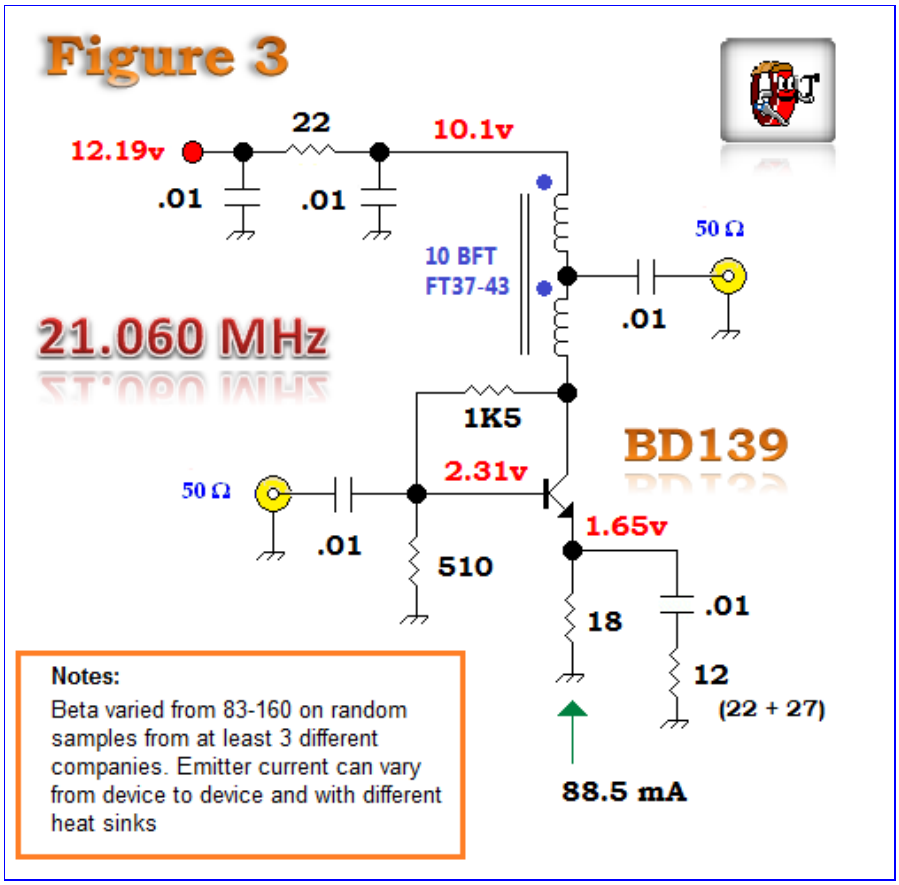
The Driver Amplifer



Above — A BD139 feedback amp breadboard. Click for the schematic. Wanting an output power of at least 22 dBm to drive my final, I built 3 varieties with both the BD139 and a 2N3866. Normally, for receiver amplifers, I focus on a strong input and output return loss and work hard to maintain linearity — but mostly keep the emitter current under 50 mA. At 21.06 MHz, the biggest *clean* signal I could muster with such a

broadband feedback amp was +16 to +18 dBm — well short of my goal. Click for a DSO screen shot of the 41.3 mA amp at 6.42 Vpp or ~20.13 dBm power; quite distorted.

I needed to try something else and saw the Figure 3 amplifier in some of Wes' work; wielding bold emitter current to boost signal handling and power.

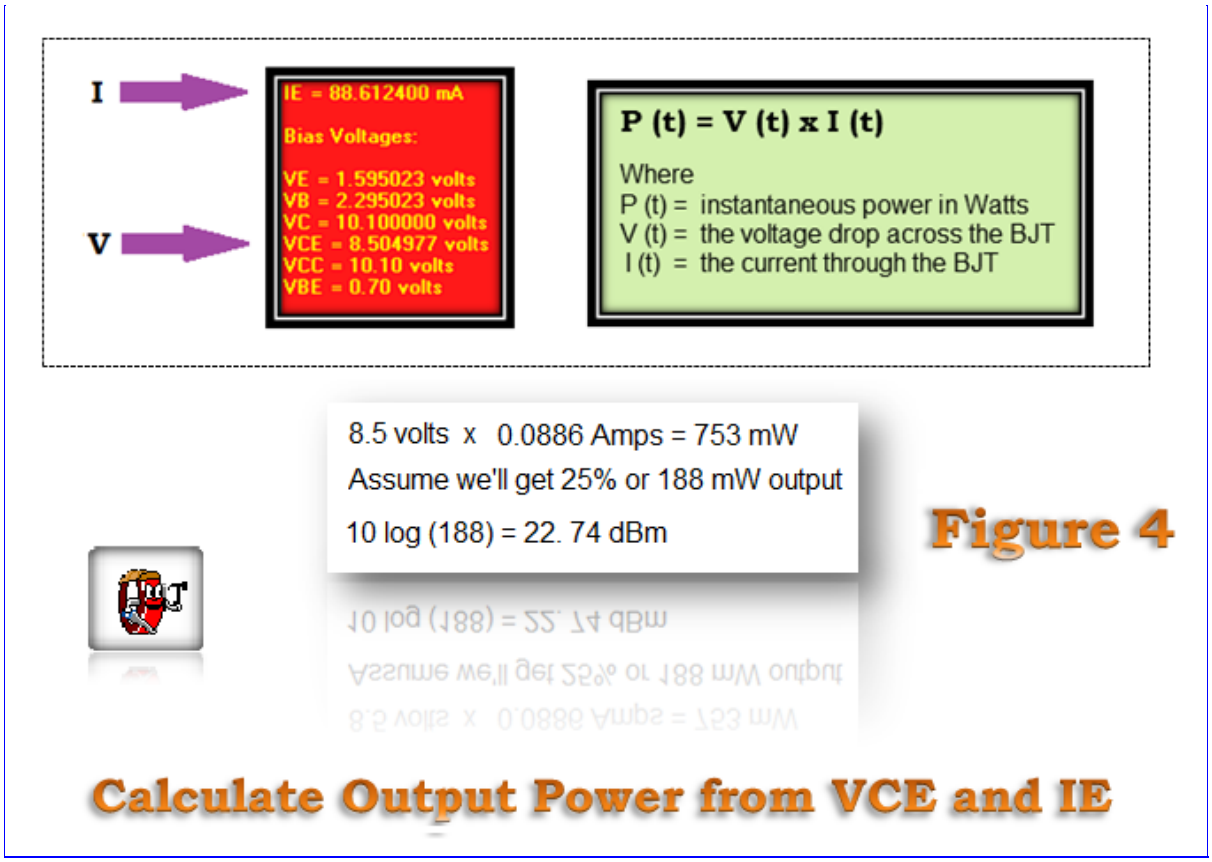


Above — The W7ZOI designed feedback amp used for a driver in all Section 1 power amp experiments. Love it! Many builders feel tempted to abandon series and/or shunt feedback and just "go for gain". My experiments suggest this approach often leads to instability within the driver circuit and possibly on down the RF chain. We should strive for linearity and stability; 2 big reasons we apply feedback. Flirt with a driver *lacking* feedback if you dare.

Click for the maximum output power attained by turning my hycas bias pot fully clockwise: 7.77 Vpp, or 21.79 dBm. You can see distortion, but since this stage will drive a Class-C transistor PA (i.e. it will switch "a diode" on and off and generate clipping), I decided it will work fine. A VHF-UHF transistor such as the 2N3866 or 2N5109 may yield as much as 3-4 dBm more power, however, you'll see more clipping at higher driver levels. I stuck with the BD139 and tried several pieces from at least 3 different manufacturers (The BD139 was made by at least 10 companies at 1 point). A Philips version purchased in 2008 yielded the 21.8 dBm, so I bolted it to a home built copper clad heat sink and soldered it in place.

Calculating Transistor Output Power Off the Bench

Design Question 2: How do I mathematically estimate my Class A amplifier output power?



Above — Estimating output power from my Figure 3 amplifier. I measured a static DC Beta of 125 as shown in Section 5 on this page. Then I calculated VCE (the voltage across the transistor) and emitter current with my program NPN-DC Bias. Wes wrote a similar and likely better program called Biasnpn08.exe that's included on the EMRFD compact disk.
You may also measure your VCE and IE directly and improve the above calculation with measurements. I do this frequently.

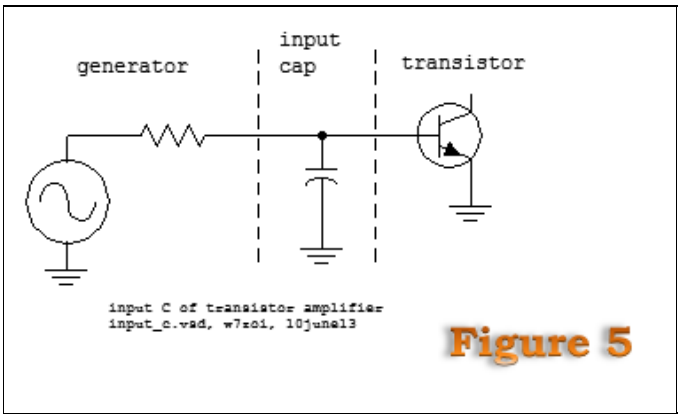
Class A amps exhibit about 25% efficiency and also imagine that Figure 3 will bear some gain compression.

My measured power = 21.8 dBm — it amazes me how close a non-bench calculation using DC analysis can get to a AC bench measurement. The Figure 4 procedure is apparently a bread and butter RF design tool.

The Power Amplifer Input

Design Question 3: How do I match my driver output to the Class-C power amp input to maximize power?

This question provided a few nights of reading and experiments; time well spent. We have 2 main ways to match our driver output to the PA input: broadband or tuned. First, lets explore the problem theoretically:



Above — A simple model sent by Wes: driving an amplifier with a power source. As amateur designers, simple models help us understand complex concepts.

A pure voltage source doesn't really exist. We don't apply a voltage to the base of the PA; rather, we apply a power source (our signal generator) — a signal voltage followed by a series impedance that's usually just a source resistance for a load. The power source + source impedance together *source* the load that includes the load capacitance represented as an input cap in Figure 5. You may wish to read about Thévenin's theorem and its dual, Norton's theorem to learn more about linear network analysis.

The load presents an impedance where the power is delivered and either dissipated in its resistance, or reflected back to the source when the load contains a reactance . At our PA input, this capacitive load reflects some of the signal back and induces a pole that corresponds to a "capacitor" time constant.

If you consider our power source, its source resistance, the PA input capacitance and just omit the effects of Q you'll see that the the voltage across the capacitor decreases as frequency increases. If you insert a 2:1 turns ratio transformer after the driver, you'll drop the AC voltage by 2 and its resistance by 4. For the same capacitance, the 3 dB cutoff frequency of this single pole filter increases. Also, the lower output impedance drops the driver amplifier voltage gain and may boost stability.

In short, we need a low driver source impedance since the Class C bipolar amplifier base-emitter junction has lots of capacitance To get current flowing into the PA base, we drop our driver output source impedance down *low*.

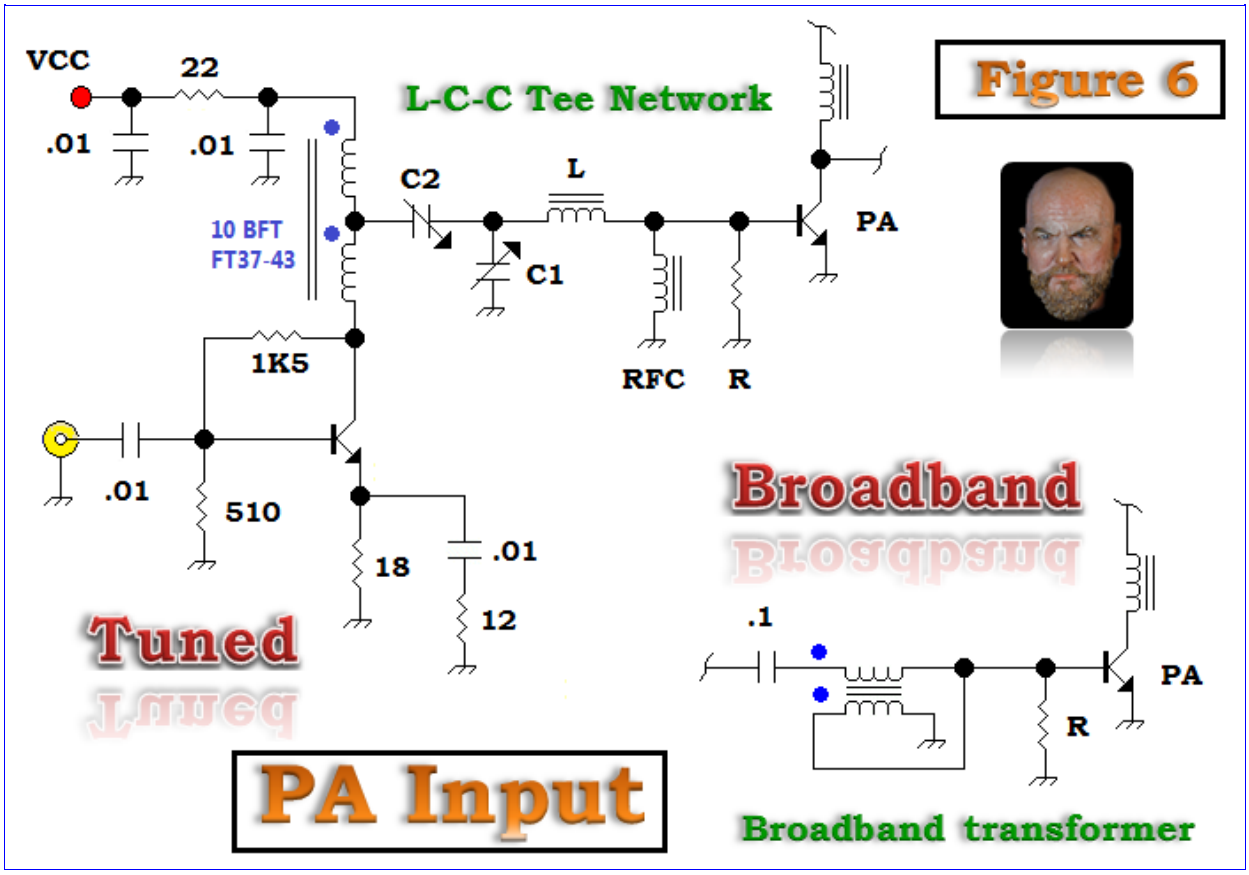
I thought — couldn't we just neutralize this capactance with an inductive output impedance? Thus our PA driver stage would drive a resistive load and everything would be "ducky". I consulted Wes, plus viewed some math and circuitry.

You can't effectively neutralize a bipolar transistor amplifier, since the capacitance isn't constant and behaves more like like a varactor diode rather than the C shown in Figure 5.. Even when forward biased, diode capacitance depends on the instantaneous voltage that appears across the diode. In contrast, 'FETs prove less troublesome and thus it's possible to achieve some neutralization with a 'FET amplifier. Please view the power stage from EMRFD Figure 11.14. Check out the push-pull pair of IRF511s. C5 and C6, at 22 pF each, are neutralizing capacitors.

In some cases, the capacitive reactance may be absorbed as part of a resonant circuit on the PA input. For example, a low-pass filter might soak up some of the C.

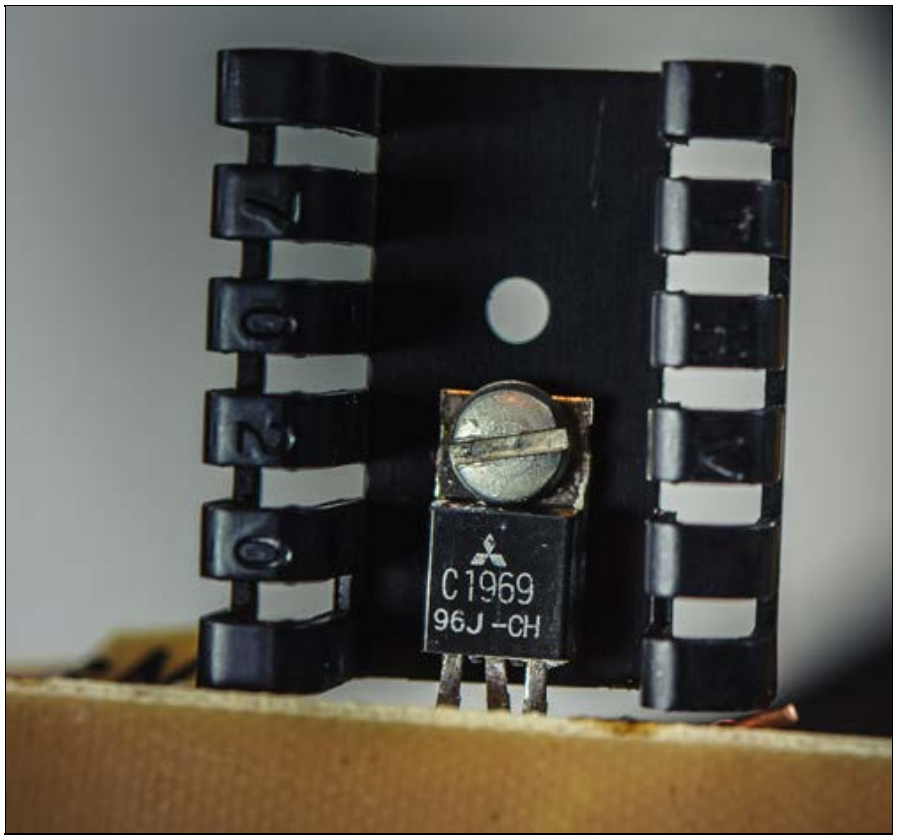
In all cases, our goal is drive in some power and switch the PA transistor solidly on and off (i.e. technically approaching class B but definitely class C because conduction will be less than 180 degrees) — a low output impedance can drive that input capacitance.

Driving final(s) with a very low output Z rings true even if your driving the latest (super-costly) ultra-low loss, high breakdown voltage, silicon GaN FETs to output >100 Watts at 1296 MHz in Class D, E, F, or Inverse Class F. Just get it done!



Above — Figure 6 shows 2 basic methods to drive the PA with a low impedance: single frequency network or broadband step-down transformer. I'll discuss both methods in detail later. Let's move away from theory and lighten things up for a couple of minutes.

Above — A short 10X probe ground lead boosts Tx measurement accuracy; in particular, for FFT analysis with your DSO. You can read or see **why** by searching the web for text and/or videos concerning oscilloscope probes.



Above — The long gone 2SC1969 works great. Beware of counterfeits however.

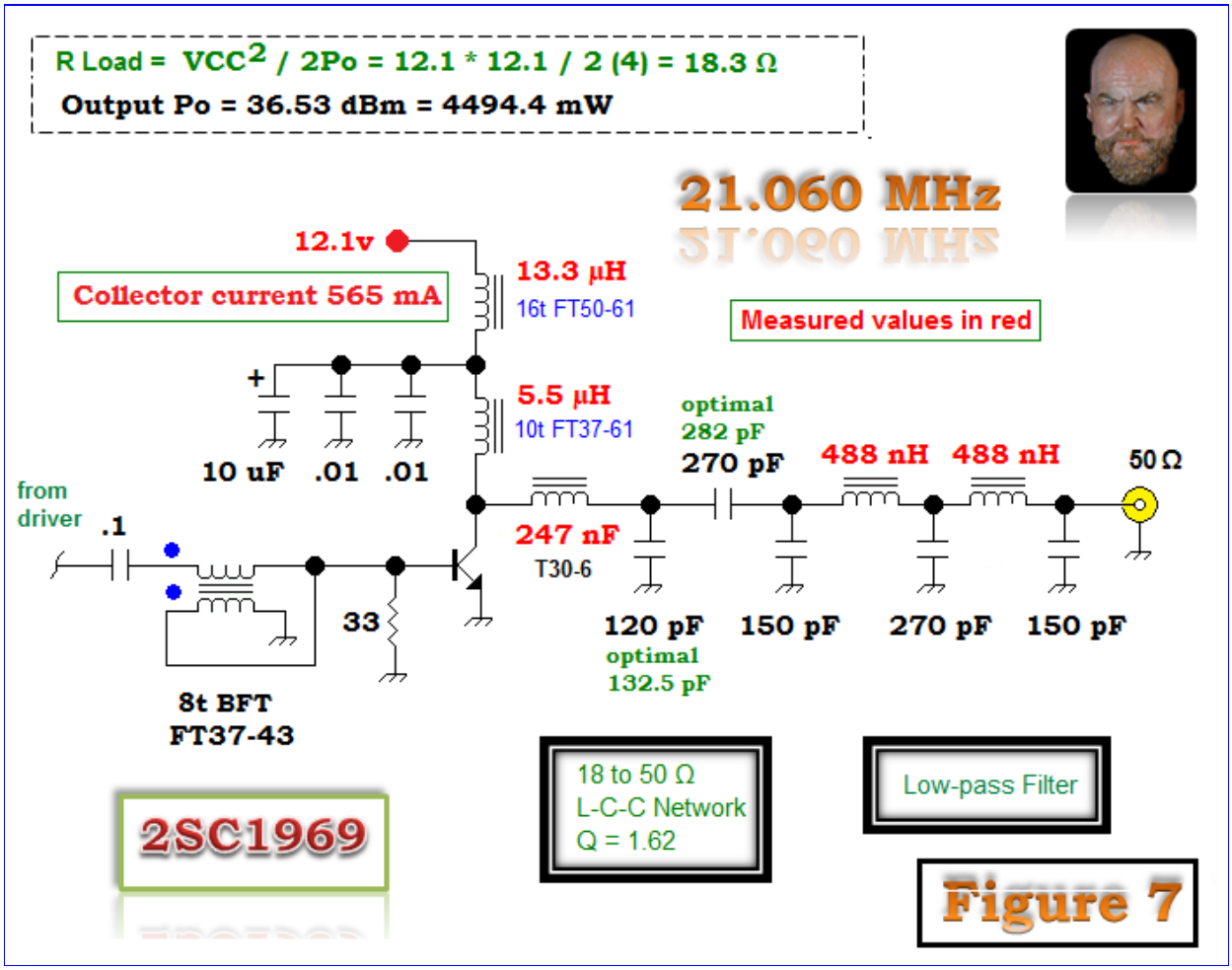
Bipolar transistor selection troubles us since most of the 'old standbys' went obsolete way back when. I've been collecting dusty, old *QRP* driver and PA transistors for decades and my bag includes BD139, MJE181, 2SC2509, 2SC1969, 2SC2166, NTE123, 2N3053, 2N4037 and 2N2219A. For VHF, apart from the 2N5109/2N3866 standards, I've kept a few MRF607, 2SC1971 and 2N3553.

VHF/UHF transistors exhibit tremendous gain at lower HF, but they might oscillate without due care.

Some rules I've gleaned: The fT of a BJT should be at least 3-5 times the operating frequency; the VCEO should be at least double the VCC (more is better) ; don't exceed the maximum heat dissipation, nor maximum current. In short, read the datasheet. I won't cover heat sinking, nor gain compensation for multi-band amplifers on this web page; for it's been done by far better authors.

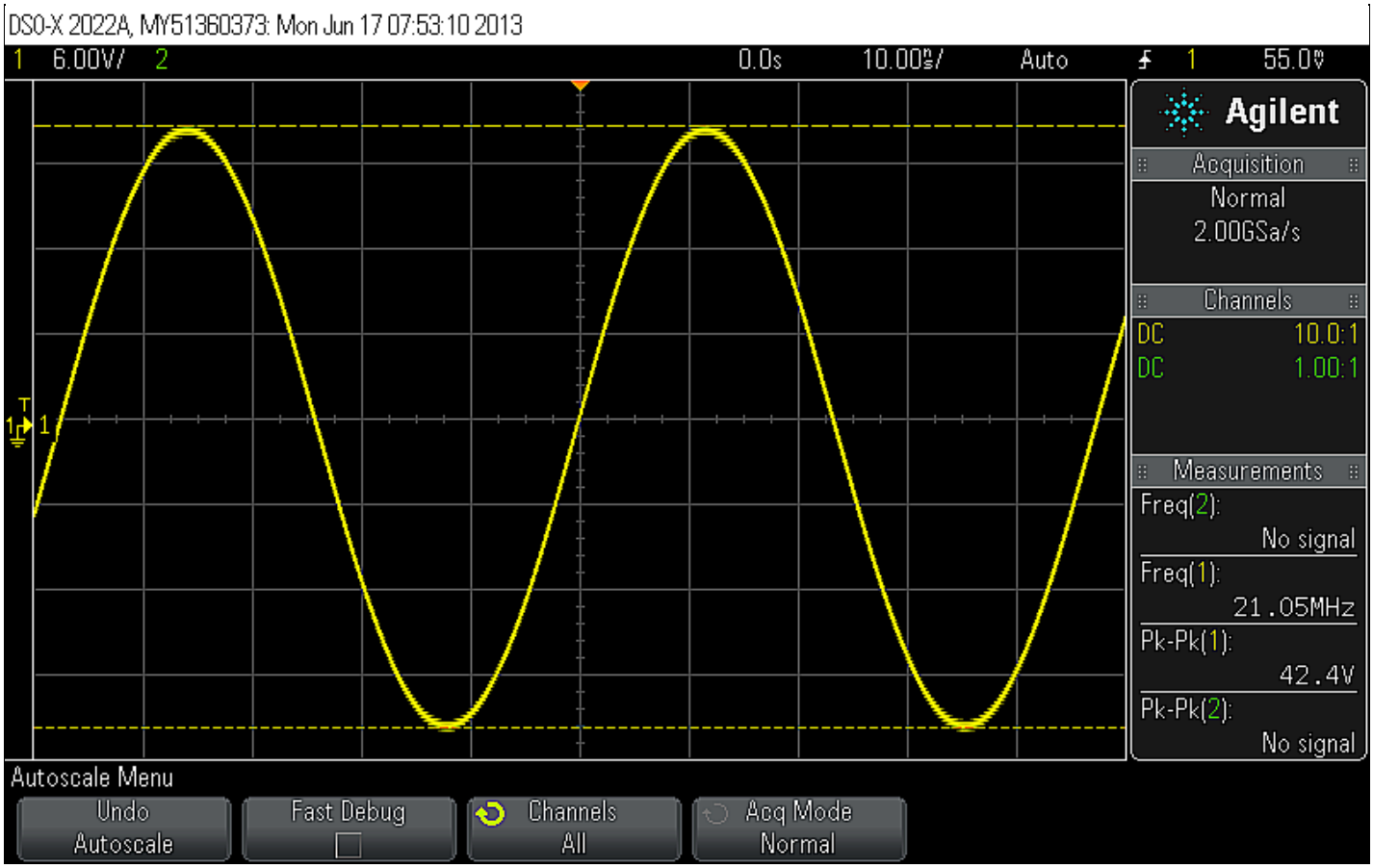
Lets make a basic Class C power amplifier based upon the RF blocks presented thus far.

Class C PA with Broadband Input and L-C-C Tuned Output



Above — Input matched PA via a broadband transformer (a bifilar wound transmission line transformer with a 4:1 Z ratio). A great example of this design includes the Figure 2 of Ugly Weekender transmitter by Wes. The 50 Ω output Z of the driver is transformed to 12.5 Ω. to drive the PA input. While a conventional link-coupled transformer works okay, the closer coupling of a twisted wire transmission line transformer often affords a better impedance transform and lower losses. You might experiment with a 9:1 Z transformer instead of the 4:1 shown. Click for a highly compressed photo of 1 of the full breadboards.

The 33 Ω resistor helps lower gain and stabilize the amp. I ran 33 to 100 Ω applying experimental methods. Read more about the PA base resistor and coil XL requirements in the text following Design Question 6.



Above — The 'scope output from Figure 7. 42.4 Vpp = 36.53 dBm = 4.49 Watts.

L-C-C Output Matching Network

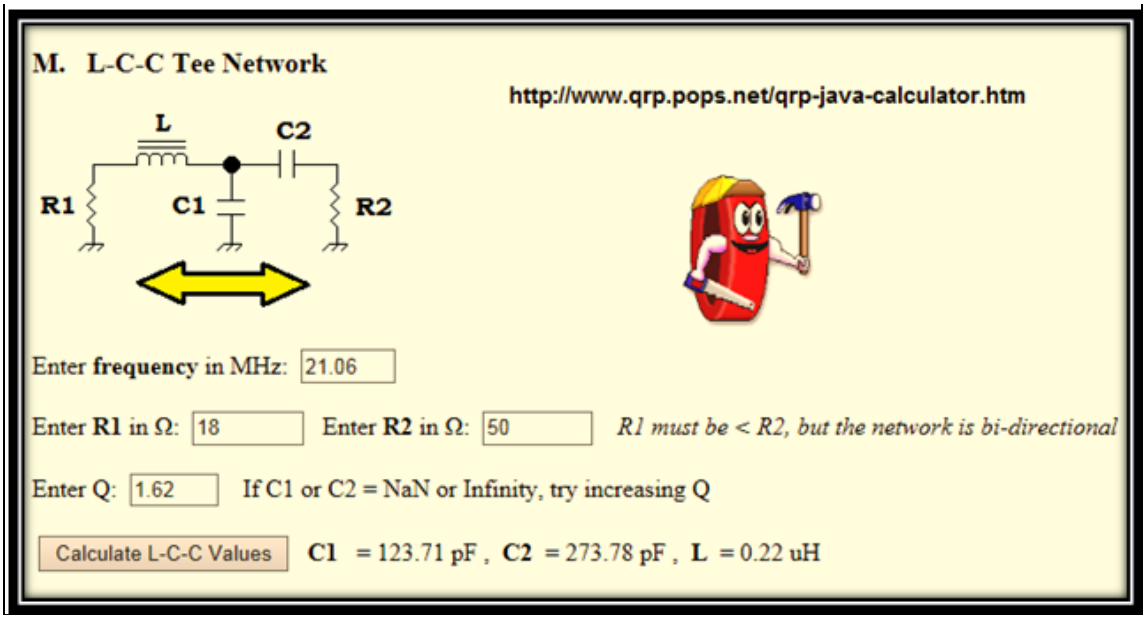
3 discussion points arise from Figure 7: [1] the output matching network. [2] the collector choke + decoupling circuitry, and [3] the low-pass filter. I'll discuss choosing a collector choke and decoupling in Section 2. We design our low-pass filters with tables or software; I used an old ARRL Radio Amateurs Handbook and then tweaked it in GPLA to get standard value capacitors. Click for a GPLA simulation of my filter. The L-C-C network also provides some low-pass filtering.

To calculate the collector output resistance, we use the formula R(load) = Vcc^2/ (2*Pout) as shown in Figure 7 for 4 Watts. For 1.5 Watts at 12 VDC --- R(load) = 12 * 12 / 2(1.5) = 48 Ω --- no output matching network is required; making 1.5 watts a popular transmitter power for new builders eager to learn about Class C.

Design Question 4: How do I make an L-C-C matching network?

Let's stick with 4W power and make an L-C-C Tee network for a 18 to 50 Ω transform. I prefer the L-C-C network over the L-network because you get 2 capacitors to tweak and I find it much easier to match a wide range of impedances compared to the L-network. Since we may pinch, or expand the inductor windings, in reality, the L-C-C gives you 2.5 possible tweaks.

Normally I wind the L on a number 6 powdered iron toroid with 26-28 gauge magnet wire in size T25, T30, T37, or T50. 26 gauge wire allows for good coil squishing and expansion with thumb and forefinger. For design cycle capacitors, I mostly solder in those horrible, low Q, orange, green, or red muRata trimmer caps such as the TZ03Z500F169B00. Although rare, ceramic-mica trimmers work well.



Above — To calculate L, C1 and C2 we may employ software to obtain starting L and C values. I recommend Wes' ZMAT08.exe, a Ladpac program that ships on the EMRFD CD. I wrote my own java-script applet with less prompting and error checking and put in on the Design Center web page as **M**. Note I used 18 Ω as the load resistance in my calculations.

During actual bench assessment of your network, output your PA to a 50 Ω dummy load rated for that power and preferably measure the peak to peak voltage with a short-grounded 10X probe.

For Figure 7, I wound and tack soldered in a measured L of about 230 nH and placed a 100 pF cap plus 6-50 pF trimmer for C1 and a 220 pF + 22 pF cap + 6-50 pF trimmer cap for C2. The tuning procedure goes like this:

Tune the capacitors for maxmum power. Try squishing the L, then re-tweak C1 and C2 and see what happens. Did the Vpp go up or down? We want it to go up, so if the Vpp went up you needed more L, and perhaps still need more. If you squish the coil again (or remove it, add 1-2 turns and then replace it) and the Vpp goes down, you increased the L too much and need to go back down. If the original Vpp goes up when you expand the coils, you had too much L and need to do the opposite procedure just described. Re-tweak C1 and C2 after every inductance change and calculate the power in dBm and/or watts on your bench.

Through trial and error you'll find the optimal L. If nothing works, suspect a bench error, or device failure. Occasionally, I add too much, or too little fixed C in parallel with the trimmer cap. Another time, my transistor was fried.

Once you've extracted the maximum power, remove the cap(s) that make up C1 and measure them. Do the same for C2. I replace C1 and C2 with the nearest equivalent standard value capacitor or (parallel capacitors) and then repeat the tuning procedure to find the optimal L for these 2 new capacitors. Normally, after optimizing the Vpp, I'll remove and measure L to confirm my L-C-C network design. You can use the L just removed or make a tidier version to permanently solder in.

Once you solder in the "keeper" L, you'll need to expand or compress the links a tad to peak your Vpp once again. Although it sounds like a chore, I love measuring a higher Vpp with each successful manipulation of the L-C-C network; it's quite a rush actually.

In the end, my optimal measured values were: L = 247 nH, C1 = 132.5 pF and C2 = 282 pF. In my final, I substituted C1 = 120 pF and C2 = 270 pF as shown in Figure 7. Although I designed for 4W, I squeezed out 4.49W with my L-C-C network.

Design Question 4: How do I choose my L-C-C matching network Q?

The easy answer is the degree of impedance transformation sets the minimum Q and thus choosing the best Q becomes the central question.

The theory regarding loaded versus unloaded Q + network losses and such may bog down many, but in context, a lower Q leads to a more efficient network with less loss. Low Q values also means we tweak larger value capacitors that minimize the effects of stray capacitance. In the collector output network of Figure 7, the Q of 1.62 gave more power than an earlier Q = 5 network. Further, during design, I sometimes slightly adjust the Q in software so I get closer to standard value caps for C1 and C2.

Don't be afraid to remove C2 and just make an L-network. Again, just experiment on your bench to gain confidence.

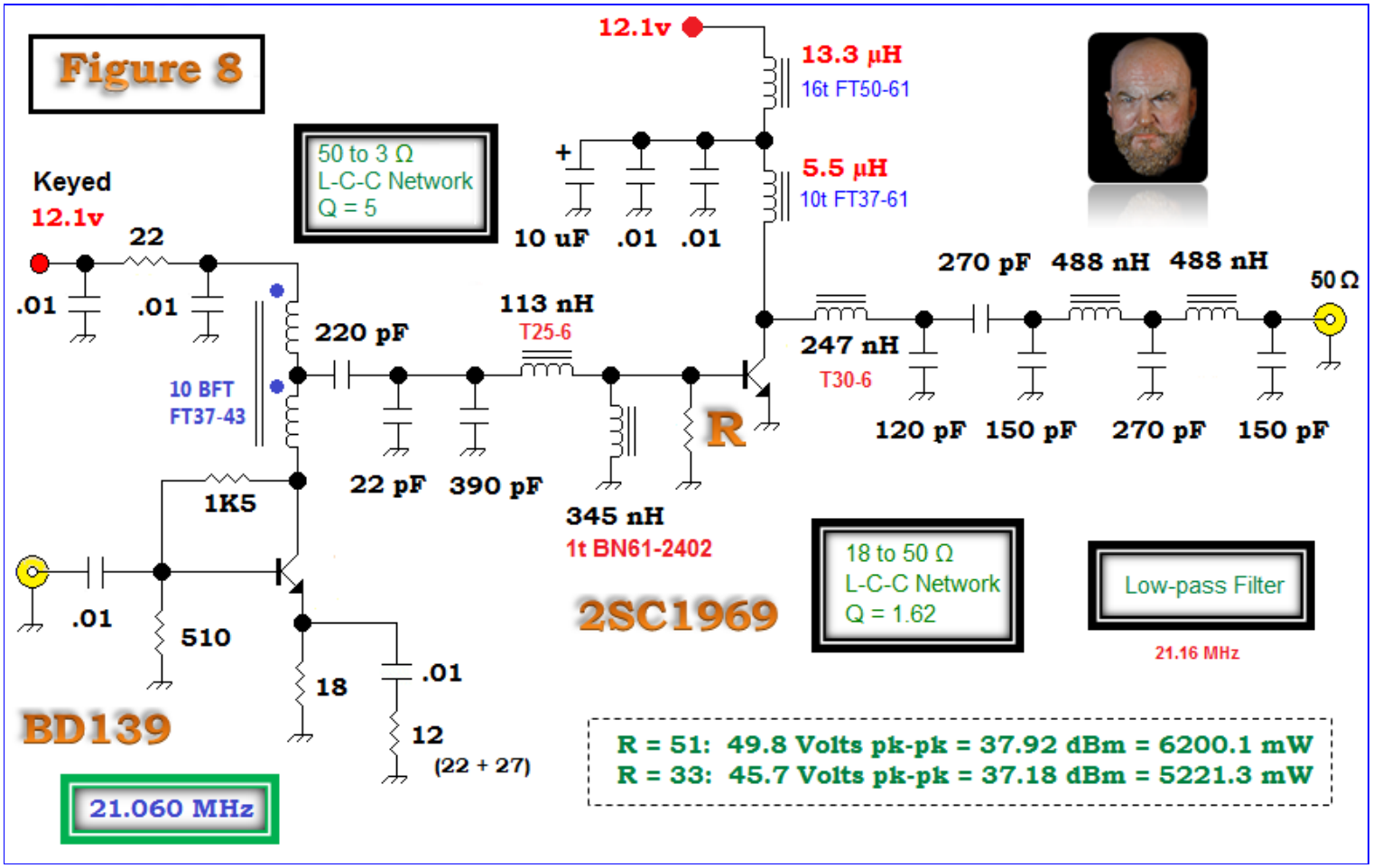
I tried making this circuit with a BD139 in the PA slot and managed to squeeze a mere 3.1 Watts power at 12.1 VDC. Others, such as Wes and Jason, NT7S delivered more power into 50 Ω at and even above 21 MHz with a BD139 PA.

Factors such as transistor specifications, frequency, VCC, and driver power may affect the output power. In the end, the 2SC1969, a final I harvested from an old Cobra CB more than 20 years ago delighted me. I constantly scrounge for old driver and PA transistors in my travels. Sadly, many old CB radios given to me over time suffered from fried finals.
You might try checking eBay, but watch out for fakes.

The BD139 grew to notoriety because it's widely available, cheap and generally does the job at or below ~28 MHz, depending on what output power your seek. I found wide variation in Beta (or hFE) and fT in the random batch I tested. Also remember we're hitting them with lots of current and they heat up and specs change. At 7 MHz, the BD139 works great. The BD139 reminds me of the MPF102 JFET; you never know what your going to get. The best I've tested were Fairchild BD139s and woefully, they're all gone.

Class C Power Amp with Broadband Input and L-C-C Tuned Output

Let's make a PA with an L-C-C Tee network tuned input: Figure 8.



Above — Figure 8. The Figure 7 PA with a tuned input network smashed my world with 6.2W power. Click for the FFT showing the second harmonic down at least 55 dB. I later confirmed this with spectrum analysis.

Design Question 5: How do I choose my PA input impedance?

PA Input Impedance

Calculating the input Z of a transistor with 100's of mA emitter challenges us.The familiar calculations like Zin = (Beta +1) x 26 /I e (mA) don't work out perfectly since our transistor is hot and Beta changes, and in my case I'm way up in fT. I asked Wes and he suggested I assume an input Z of 3 Ω (with about 200 pF of parallel C). Thus , I chose a 50:3 Ω transform to design my L-C-C network.

From my experiments, bench tweaking an L-network, or L-C-C network will accommodate anything even close to 3 Ω input impedance — 3 Ω seems quite actionable. Tune your L-C-C network in the manner described for the PA output match. Since your transforming 50 Ω to 3 Ω, the minimum Q will rise.

Design Question 6: How do I choose my PA input choke and resistor when fed with an L-C-C network?

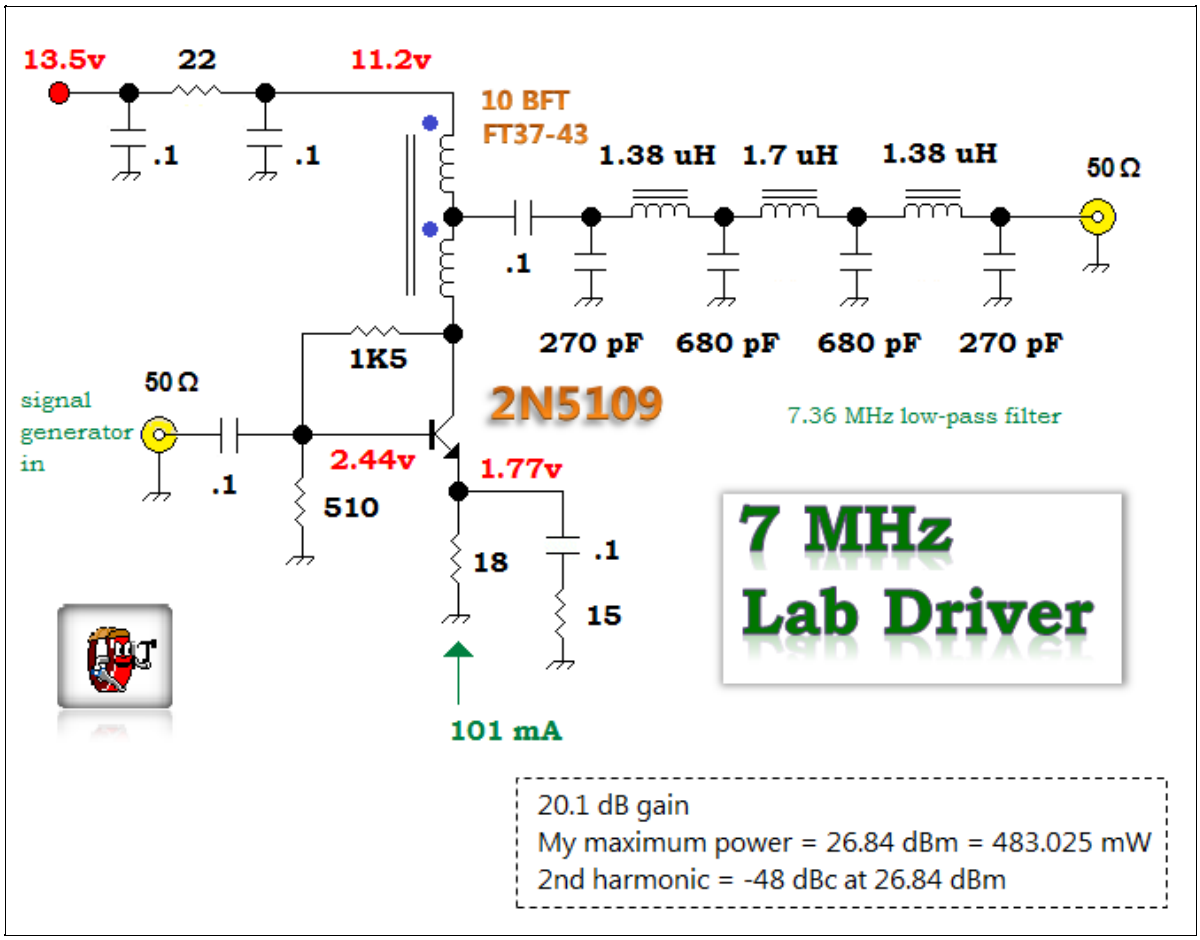
The RFC serves as a DC return path for the PA base just like in the transformer in the wideband version. We normally apply a rule that the minimum inductive reactance of a choke should be 10x the impedance it's connected to, so that's 30 Ω XL.
Since F = 21.06 MHz, the minimum choke value will be 30 Ω / 21.06 MHz x 6.28) = 227 nH. In order to keep the SRF up and for ease, I ran a single turn of 26 gauge wire through a BN61-2402 binocular core and measured 345 nH. Good enough. The # 61 mix will exhibit some Q. Likely # 43 material will work fine too.

The resistor labelled R serves to reduce the Q of *my* particular RFC, lowers gain and boosts stability. With R = 33 - 51 Ω, the amp held stable, but if I tried to increase R above 51 Ω spasm erupted. The resistor affected output power as shown in Figure 8. Again the resistor should minimally be about 10X the input Z or so. I tried an R of 10 Ω and the output power dropped down to a little under 3W as more of my driver output just flowed through the resistor to ground.

To conclude, I executed the time honored tradition of choosing parts and circuits to extract power from a crystal oscillator-driven analog RF amplifer chain. Much of this information extends to linear operation (Class A, AB) and general RF design. Have fun designing your own transmitters and as always, apply experimental methods!

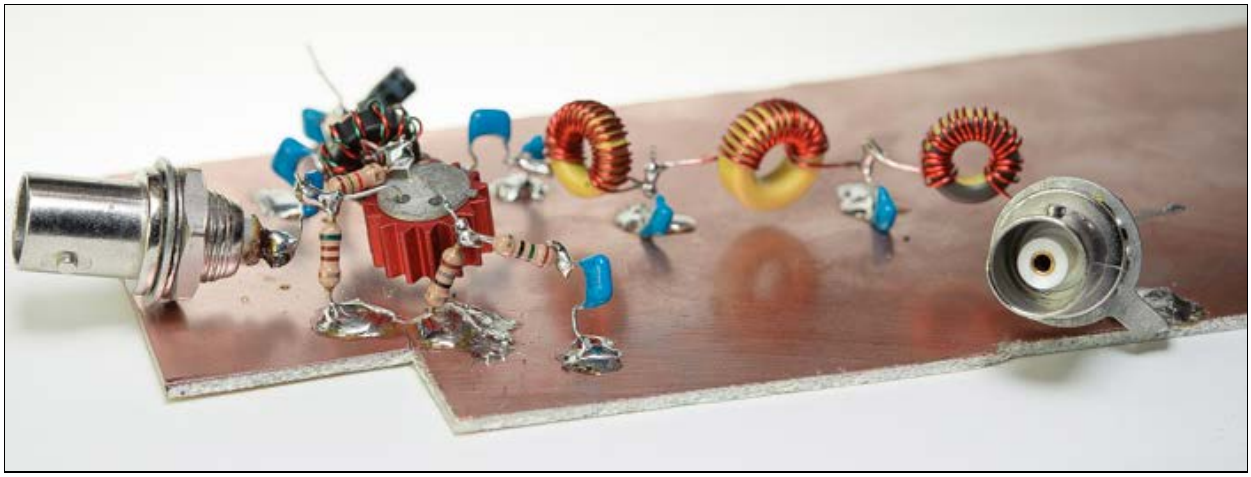
Section 2: PA Measurement with Examples

In this section, I posted some PA measurement ideas + experiments and comments about PA collector chokes.

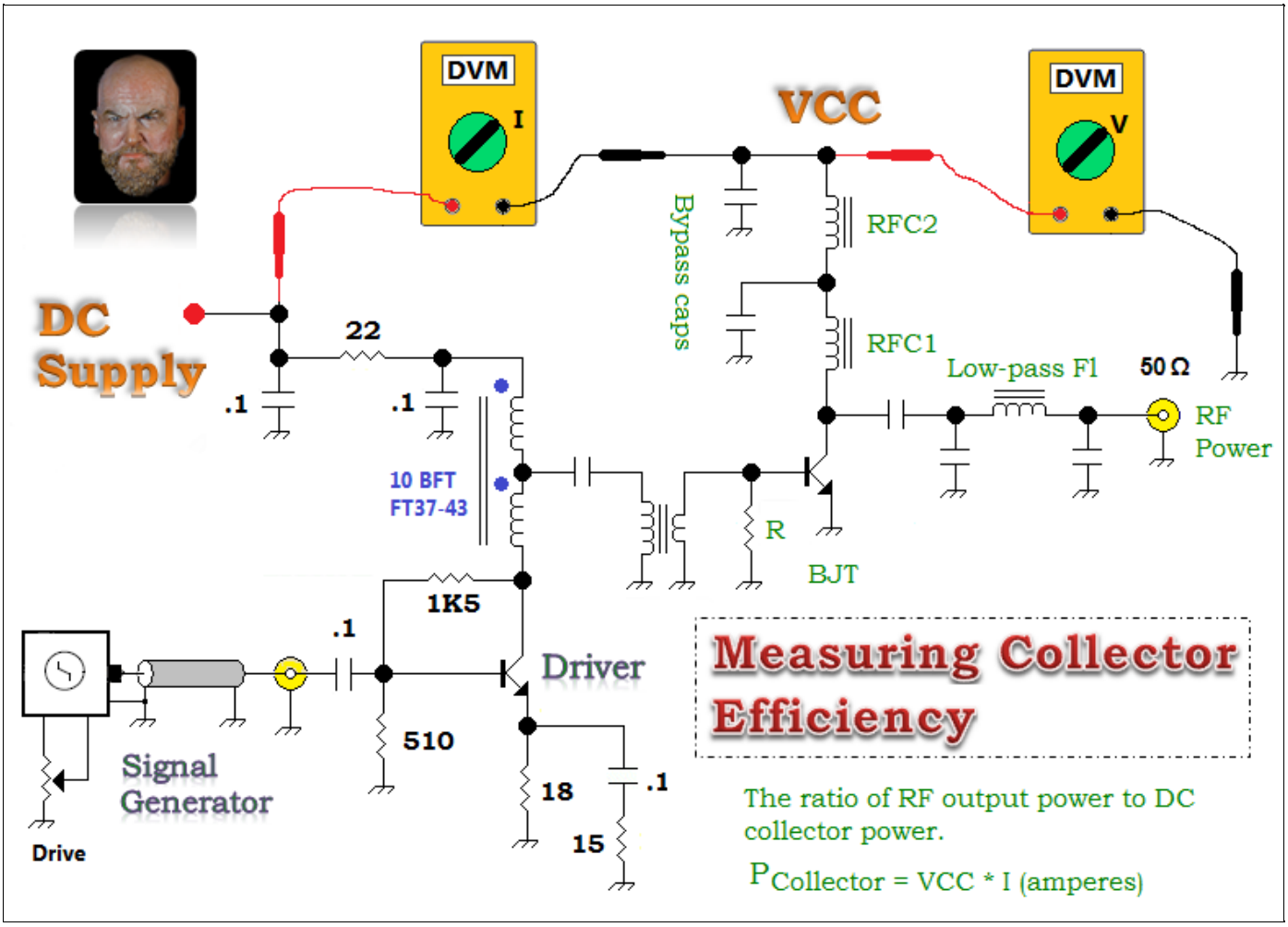


Above — To boost my signal generator output for driving *any* experimental PA up to about ~20W output, I built a high-current, sine wave driver. I excite this amplifer with my variable gain homebrew HF signal generator — in the version shown above, a 7 MHz low-pass filter is soldered on the amplifier breadboard. I'll stick this amp in a metal box sans filter and then apply the correct bench module low-pass filter via a BNC connector. Click for a photo of 3 of my HF low-pass filter bench modules.

Normally the low-pass filter output is pad attenuated to buffer and reduce the output voltage since it's rare I need up to 483 mW drive on the QRP workbench.



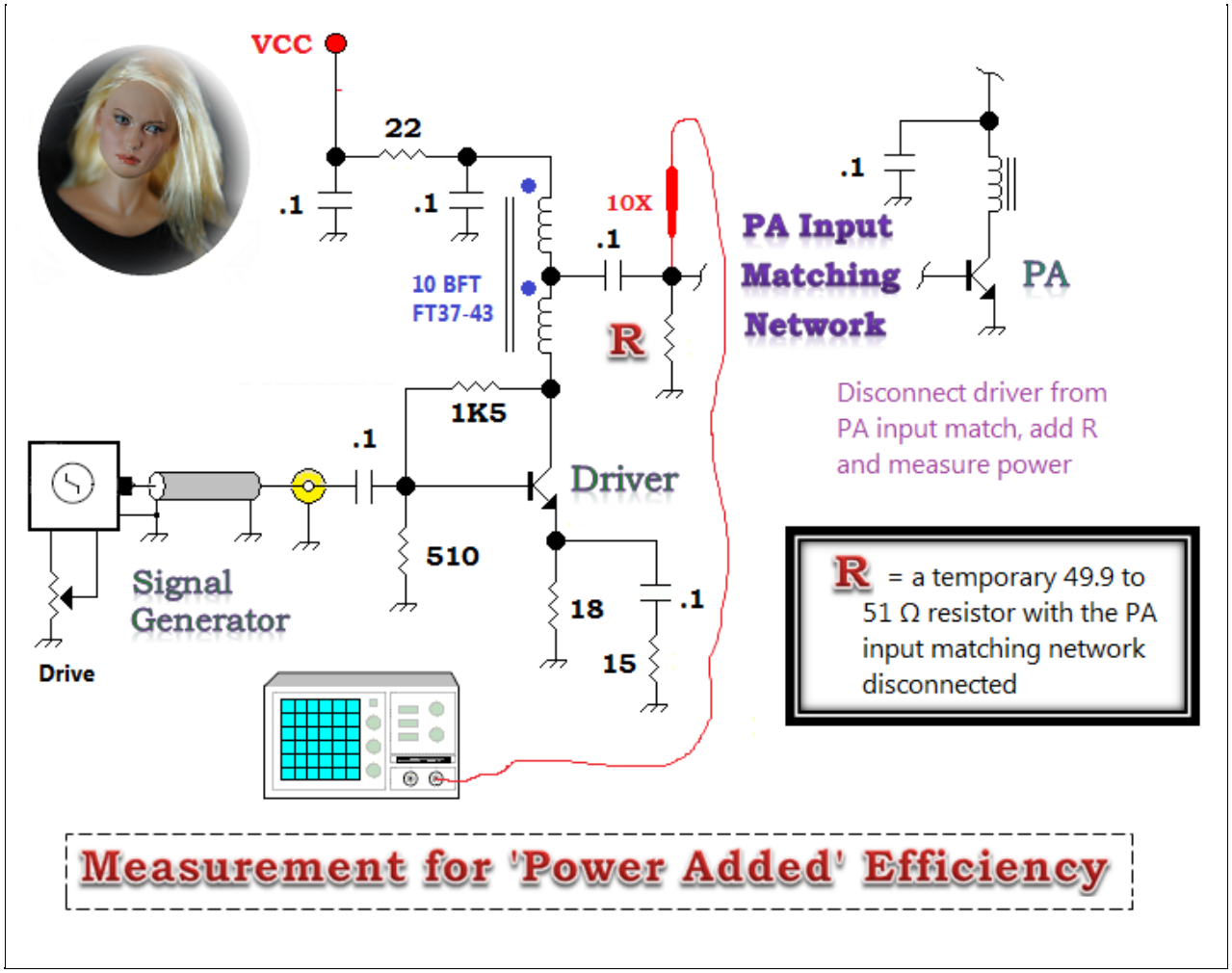
Above — A photo of the PA driver amplifer breadboard.



Above — A method to simultaneously measure collector **V** and **I** to calculate DC collector power. Two DVMs boost measurement fidelity since the current measuring ammeter's internal resistance will drop the DC supply ~ 0.2 to 0.5 volts depending on the device and current. Collector efficiency = RF output power / (V * I) .

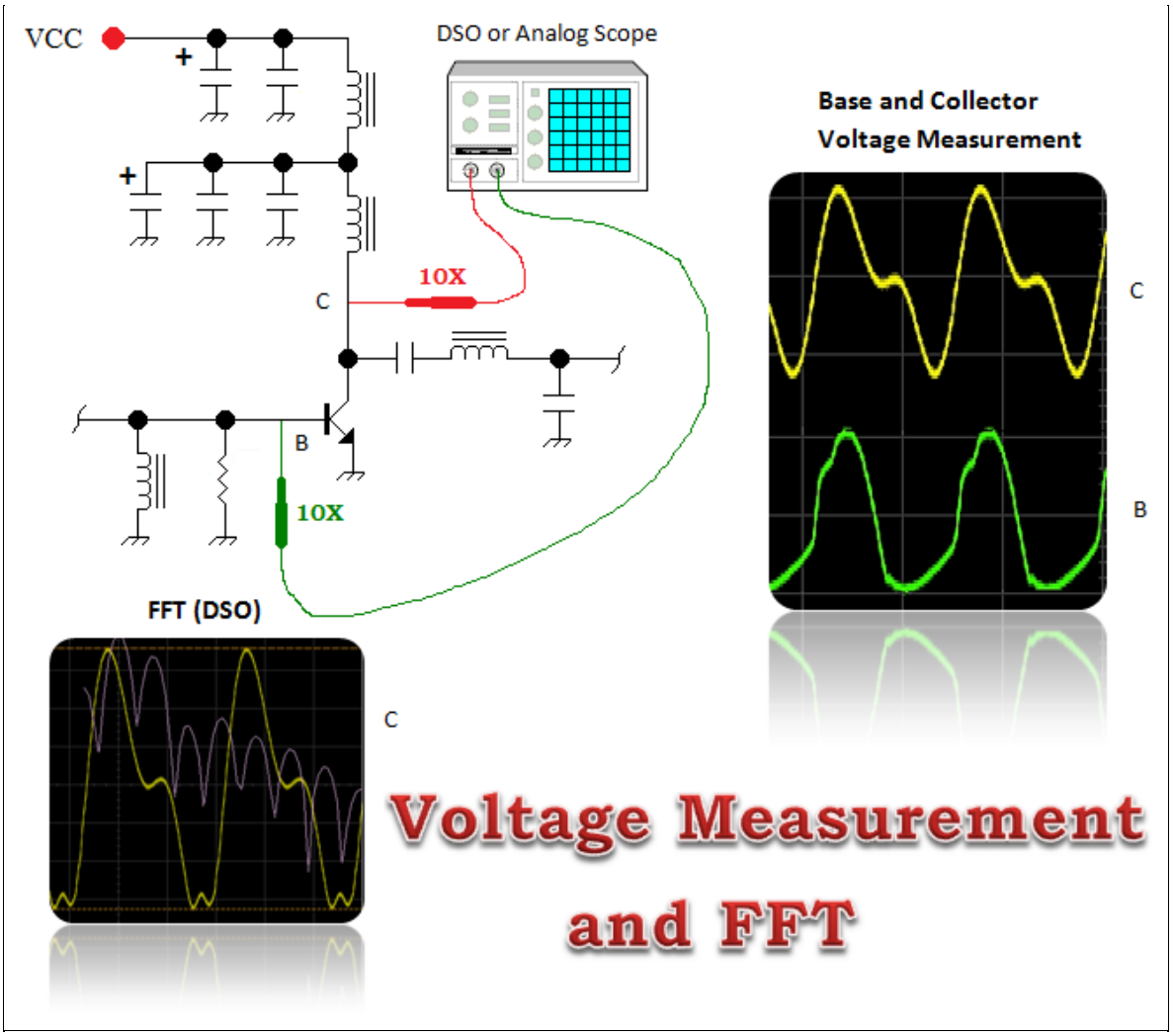
This method provides a great excuse to purchase another multimeter.

Measure VCC where shown since the decoupling and bypass filters help prevent the AC signal from affecting the DC measurement. We only touch our 10X 'scope probe on the collector. Exert due care and wear eye protection in case you blow up the transistor and create shrapnel.



Above — Power Added Efficiency considers the base drive power in addition to the DC collector power. The formula = RF output power / (DC Power + RF drive power). Although uncommon, measuring drive power helps you to assess and compare both input matching and PA efficiency metrics.

At the very least — I like knowing my PA drive power and so employ driver amplifiers with 50 Ω output impedance to allow this power measurement. I often permanently solder a 49.9 or 51 Ω resistor to the breadboard ground plane and temporarily connect this R ; plus disconnect the input matching circuit to measure driver power when I've finished tweaking my collector to antenna port circuitry.

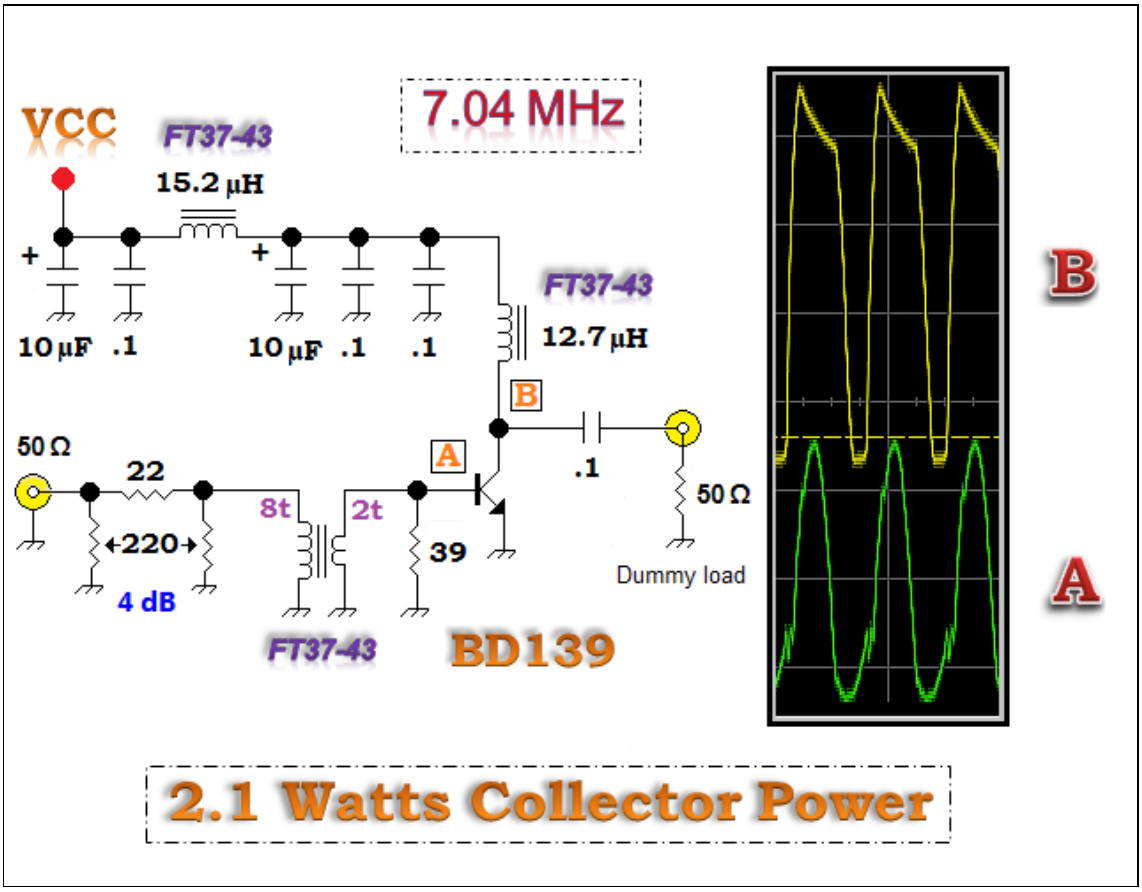


Above — Oscillocope measurements of the base and collector voltages. Most of us enjoy simultaneously running 2 channels on our 'scopes. With a good quality DSO, FFT analysis of the collector waveform may yield insight, however, this knowledge might also confound us since interpreting our measures is fully half of the task.

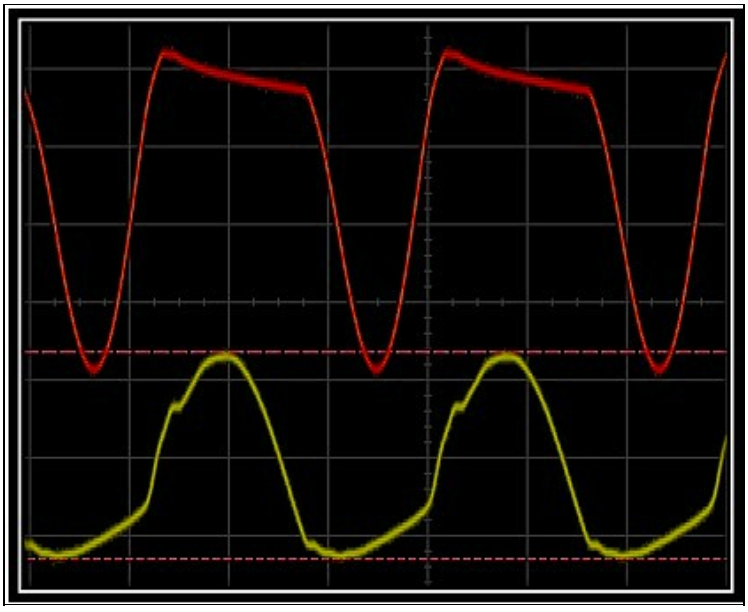
I often struggle to properly interpret measures. Some builders blog about or email me that they enjoy an innate gift at understanding electronic circuitry — I don't. I guess I lack the knack? For some of us, problems and complications block progress, but more often than not, we eventually cross the finish line.

Above — 1.5 Watt, Class C breadboard used for the experiments that follow. I changed the collector choke (RFC1) and applied various filters to examine some of the measurement techniques aforementioned. Click for a bigger photo.

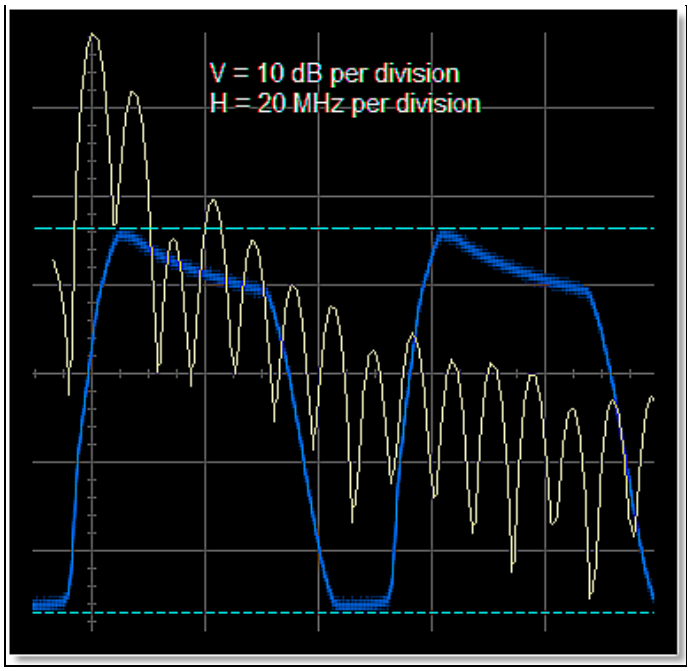
The Collector Choke



Above — A Class C amplifer with it's collector AC coupled to a 50 Ω resistor load. *No filter.*



Above — Base and collector waveforms. The unfiltered collector waveform (top) rich in even and odd harmonics ~resembles a square wave. The maximum collector voltage occurs when the current is 0 and in concert, the maximum collector current flows where collector voltage falls to near 0. As the BJT shuts off, the collector circuit kicks and rings so that the DC collector voltage may rise as high as twice VCC.



Above — An FFT of the unfiltered collector showing F1, F2, F3, F4, F5, etc. Distortion city!

The Collector Choke

I've read web pages suggesting we might boost PA efficiency by decreasing the collector choke inductance. While that's possible, in reality, it's not that simple. Experiments involving factors like collector load resistance, angle of conduction, RFC inductance/Q, matching networks, base drive configuration and output filtration are ripe with complicated math and theory that I struggle to understand — entire books and research papers explore these topics as author(s) configure their PA into Class D, E, etc. to boost efficiency. The published design work and PA devices out today are truly breath taking.

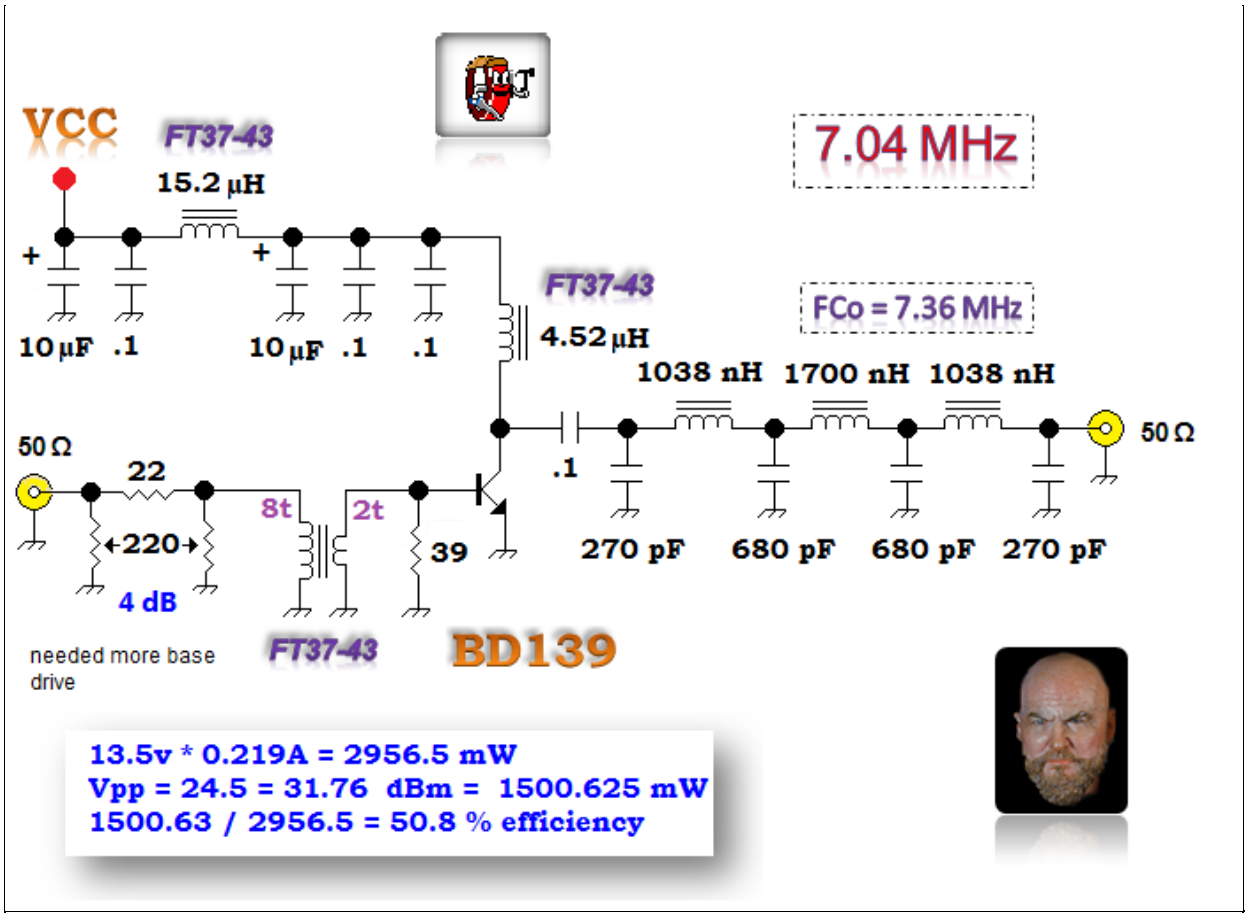
We should likely think of the BJT, collector current + voltage, the RFC, base drive and output networks as a system of parts that interact — changing 1 part might change collector current, impedance, efficiency, harmonics, etcetera. As such, the experiments that follow do not construe science; only some experiments by a neophyte to ponder and build on.

The ideal RFC passes only DC — i.e. exhibits infinite impedance to AC, but a short circuit to DC. The primary purpose of the choke is to block our high-power, distorted AC signal from flowing on our DC power lines. Some authors suggest that the minimum inductance for the inductor should be 5-10X the collector resistance and I'll leave that up to you. At HF, I wind my RFCs on #43 or #61 material ferrites while keeping the turns number low as I remember that the coil will exhibit a self resonant frequency that might cause problems if it lies near a frequency I need to filter.

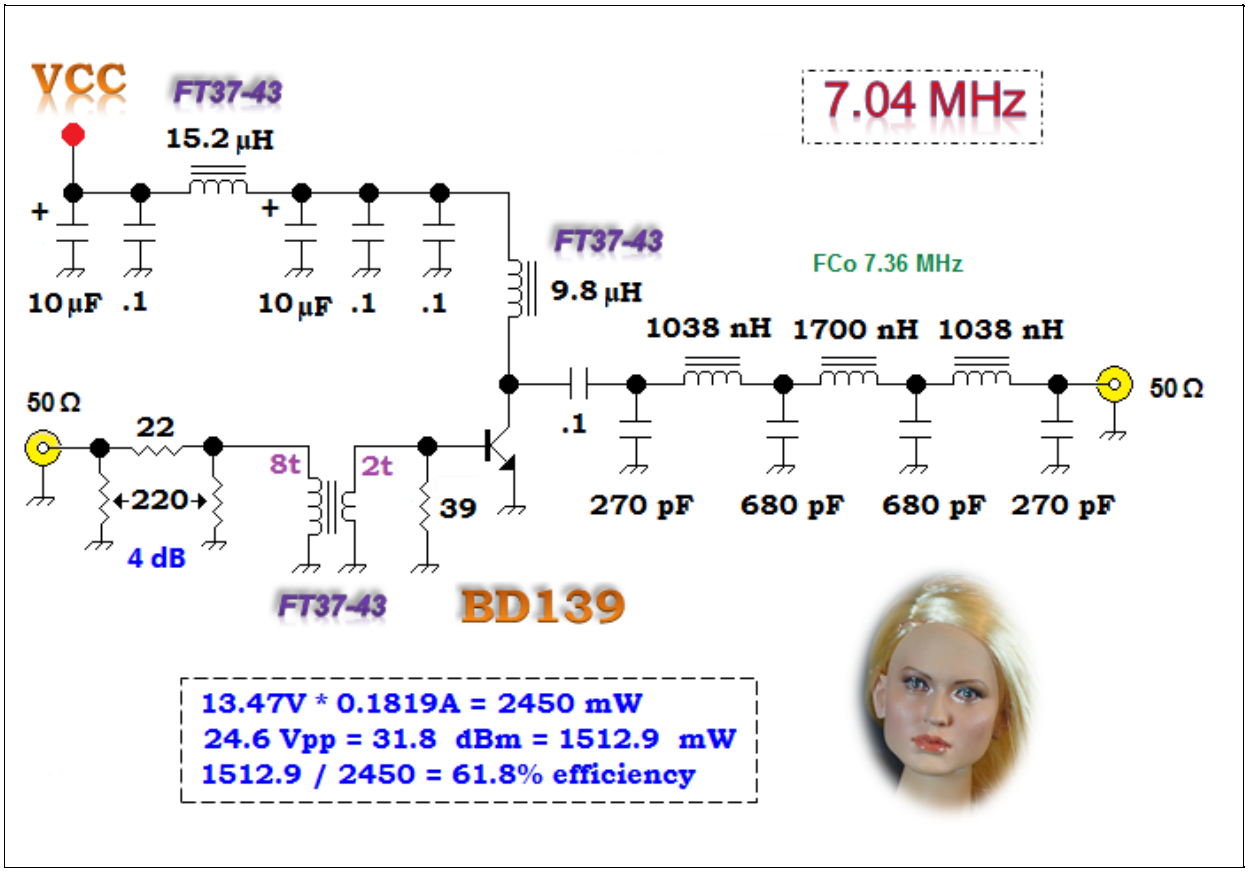
1.5 Watt Amp with Three Different Collector Chokes

The 1.5 Watt obviates the need for a matching network since the collector load resistance = 48 Ω with a VCC of 12v. I rebelled and ran 13.5 V supply for a calculated 60.8 Ω collector load R. I probably should have this matched this collector load to 50 Ω at 13.5v, but wanted a "real world" amp without a matching network. At the very least, these amplifiers illustrate the PA measurement techniques shown above. I wound the RFC on a common FT37-43.

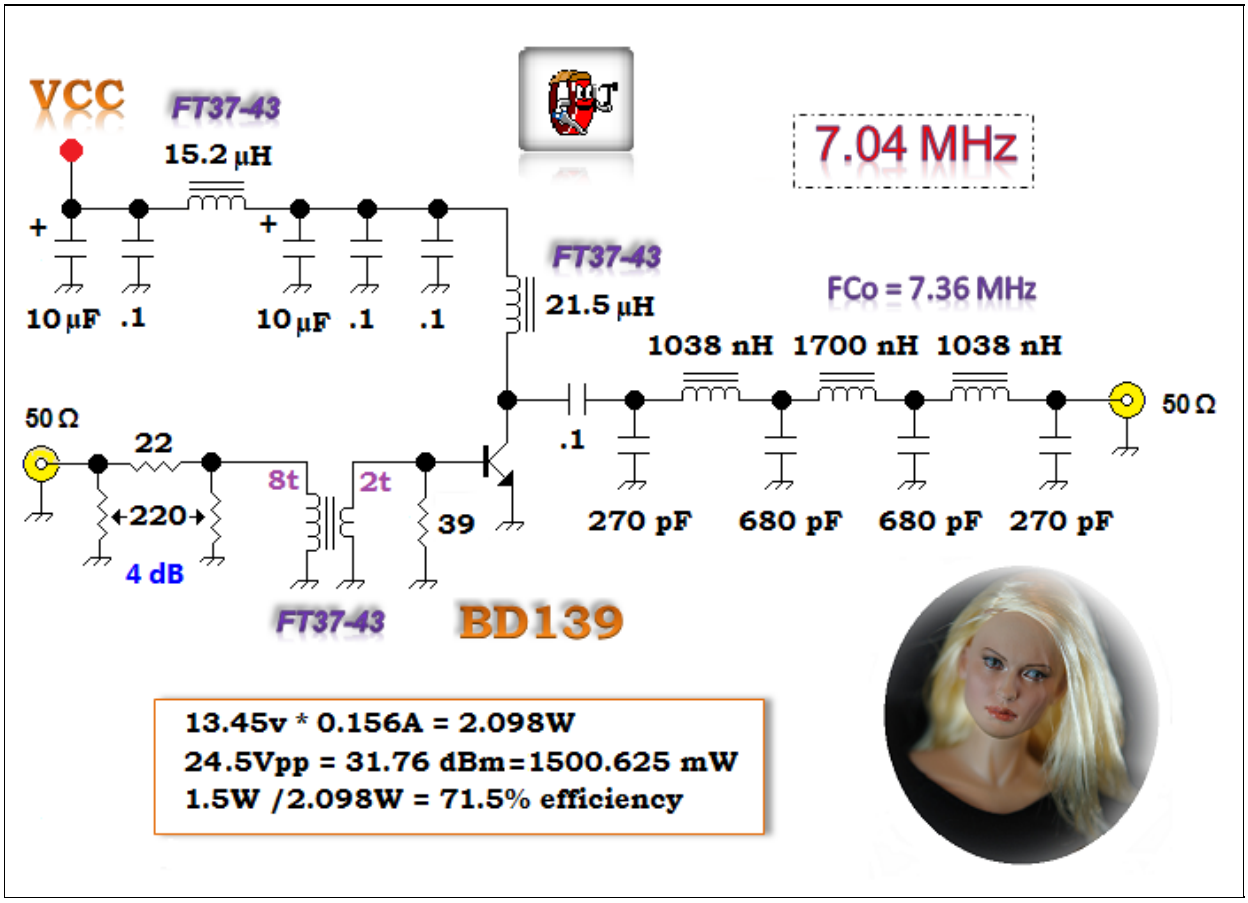
In all cases, I adjusted the base drive so that the sine wave PA output read ~24.5 Vpp while measuring collector V and I. I made DSO measures of the the base and collector voltage and collector FFTs, although forgot to save some of them as screen captures.



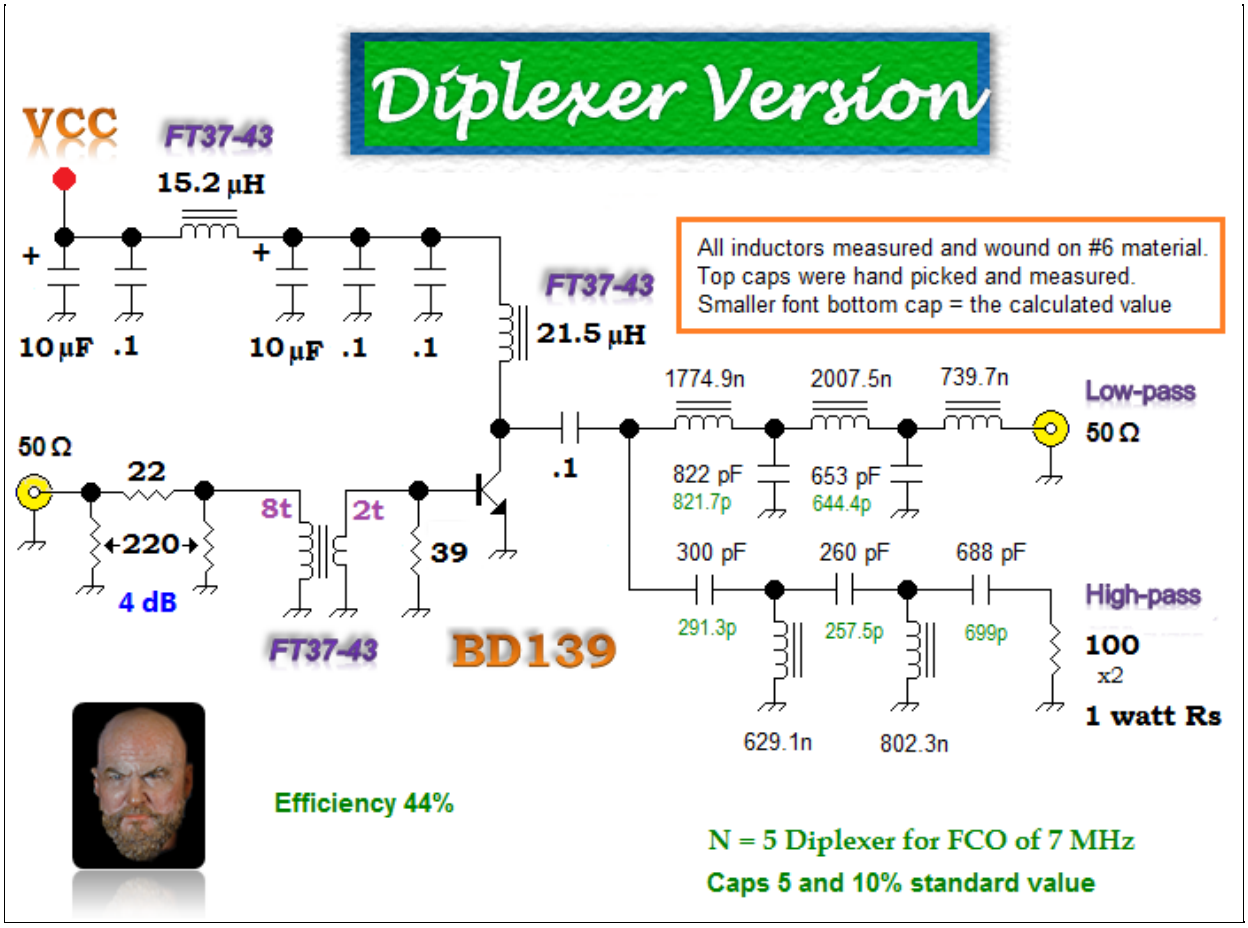
Above — The collector choke = 4.52 µH or an XL of ~ 200 Ω @ 7.040 MHz. My measurements indicate an efficiency of 50.8%. With a choke XL of only ~ 200 Ω, some of my collector AC might be flowing to ground through the bypass caps and that'll kill efficiency. Click for an FFT of the collector. This circuit required more base drive than the other 2.



Above — The collector choke = 9.8 µH = an XL of ~ 434 Ω. Efficiency now 61.8%. Base and collector tracing. Collector FFT.



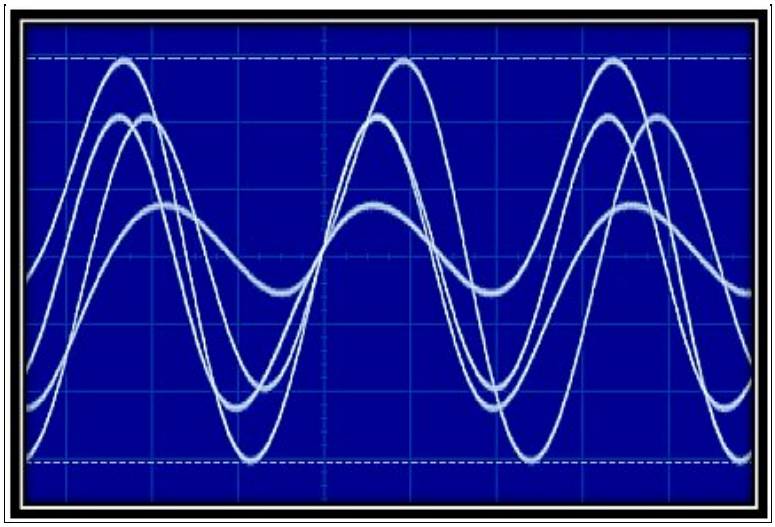
Above — The collector choke = 20.5 µH = an XL of ~ 951 Ω. 71.5% efficiency. Base + collector. Collector FFT. Sine wave FFT. To date, 71.5% is the best Class C efficiency I've measured.



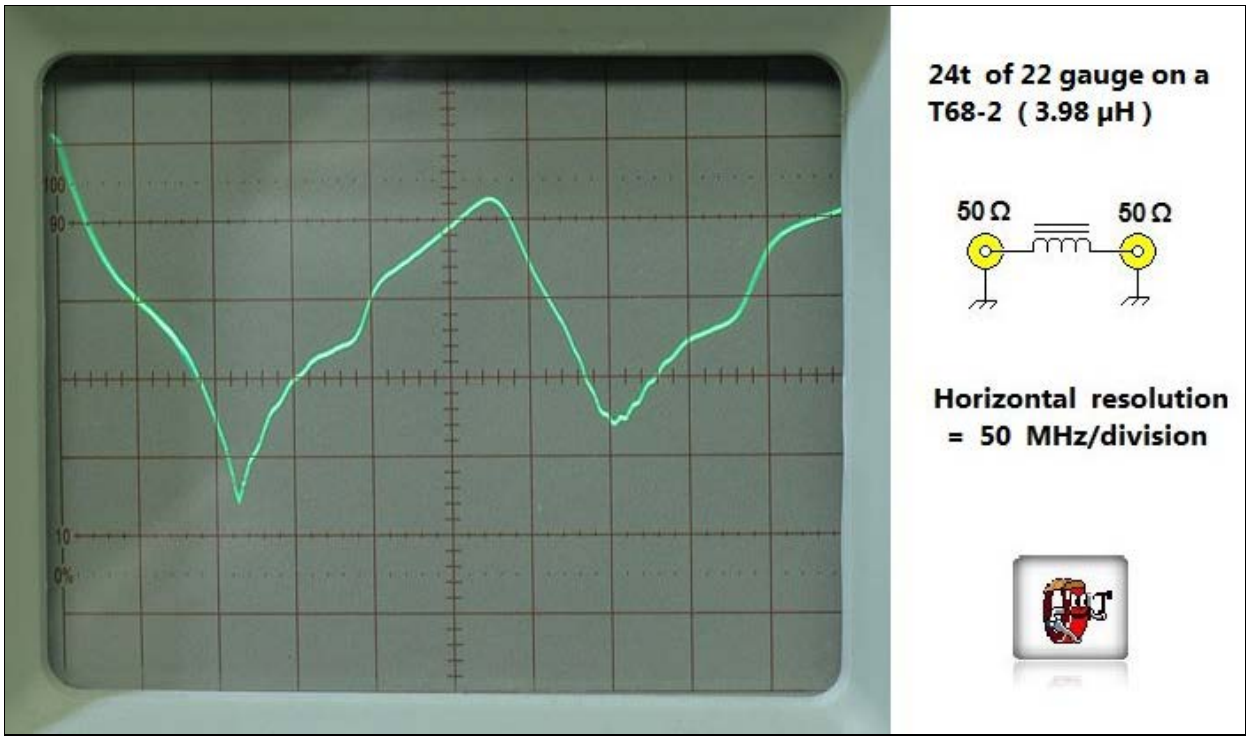
Above — A PA terminated in a N= 5 diplexer for 7 MHz. Click for the FFT of the collector. The collector waveform looks telling — resembling the 50 Ω resistor terminated version shown earlier. Unshown; I lowered the signal generator to 6.99 MHz and drove it to a sinusoidal 1825 mW to obtain a true 50 Ω collector load resistance plus get the greatest low-pass output signal and efficiency (44%) . Although diplexers will 50 Ω terminate the collector at all frequencies, they don't afford high collector efficiency.

Rogues Gallery

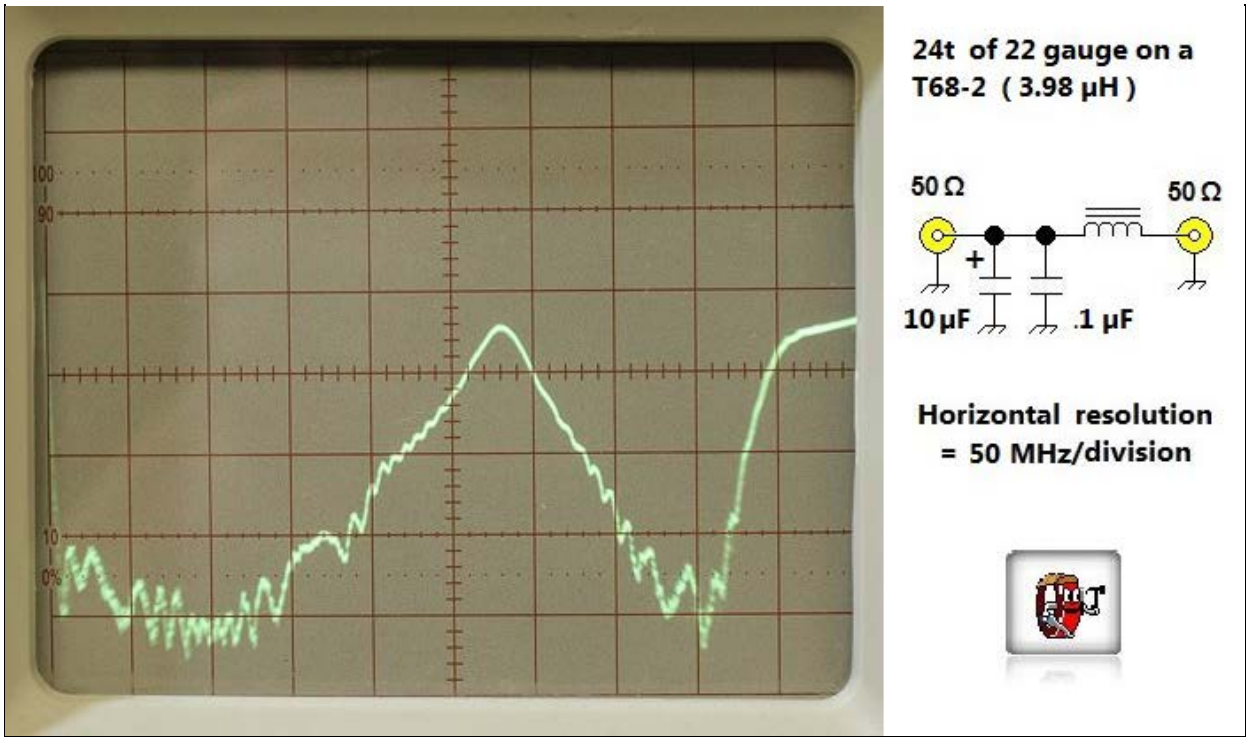
Above — Some 5W CW amplifier collector waveforms as I tried to boost efficiency by manipulating a matching network to drive up the ringing; all failures. These seemingly simple Class C PA's aren't so simple after all. Although failures — I felt thrilled by these experiments and vow to get better at it.



Above — A sine wave output of a very unstable PA.



Above — To observe a renegade series resonant frequency, I wound a choke for a 21 MHz amplifer on a T68-2 powdered iron toroid: a casual 24 turns of wire. Sweeping it shows 2 SRFs at roughly 115 and 350 MHz. 2 SRFs jarred me.



Above — I placed a couple of caps on the cold side like we normally do for our PA chokes. Likely I should have used a 0.01 µF cap instead of the 0.1 µF shown for 21 MHz, but the result would be the same: 2 SRFs remain and poor wideband bypass arises. We clearly need to consider SRF in our chokes and how it might affect our PA function and DC line filtering.

Conclusion to PA Measurement

With an oscilloscope and 2 DVMs you can easily explore boosting your Class C PA efficiency. I've seen precious little info on this topic in the amateur RF experimenter literature. 1 reputable reference = a section in EMRFD called **A Look at Some High Efficiency Amplifers**. For now, I'll temporize. Likely you'll advance faster than me, since I'm more a receiver guy and don't devote enough time to study transmitter design.

The math and concepts of efficient PA design within Class C - E flummox me. Most have just switched to driving FETs with a square wave built up from parallel inverters (high-speed logic gates in the now obsolete 74AC series) — but hopefully our experiments tweaking BJTs in Class C will carry us forward into other Tx modes.

Measuriing collector power with 2 DVMS while watching the sine wave output in your 'scope gives you incredible feedback: you can even observe thermal runaway. After purposely advancing the base drive to "heat up" a BJT, I saw the Vpp progressively drop in the 'scope, the DC voltmeter drop somewhat and then the DC current meter 'take off' : 400, 500, 890, 1200 mA and suddenly, poof, the BJT was on fire, exploded and flung 2 leads into the air. Normally, they just smoke and stink up your lab. I put a picture of a couple of blown BD139s in the photo section below. Expect to burn up some BJTs during experiments and please work safely.

Section 4: Class E Amplifier

This section unposted.

Section 5: References

1. Solid State Design for the Radio Amateur by Wes Hayward, W7ZOI and Doug DeMaw W1FB. American Radio Relay League 1986.

2. An Updated Universal QRP Transmitter by Wes Hayward, W7ZOI. American Radio Relay League: QST for April 2006.

3. Class E Amplifiers Part 1: Class E Basics. Dan Tayloe, N7VE / Ozarkcon Class E Presentation, 1994. *Search for it online*

4. The Handiman's Guide to MOSFET "Switched Mode" Amplifiers by Paul Harden, NA5N Part 1 Introduction to Class C,D,E and F. First Published in the journal "QRPp". Paul's Site

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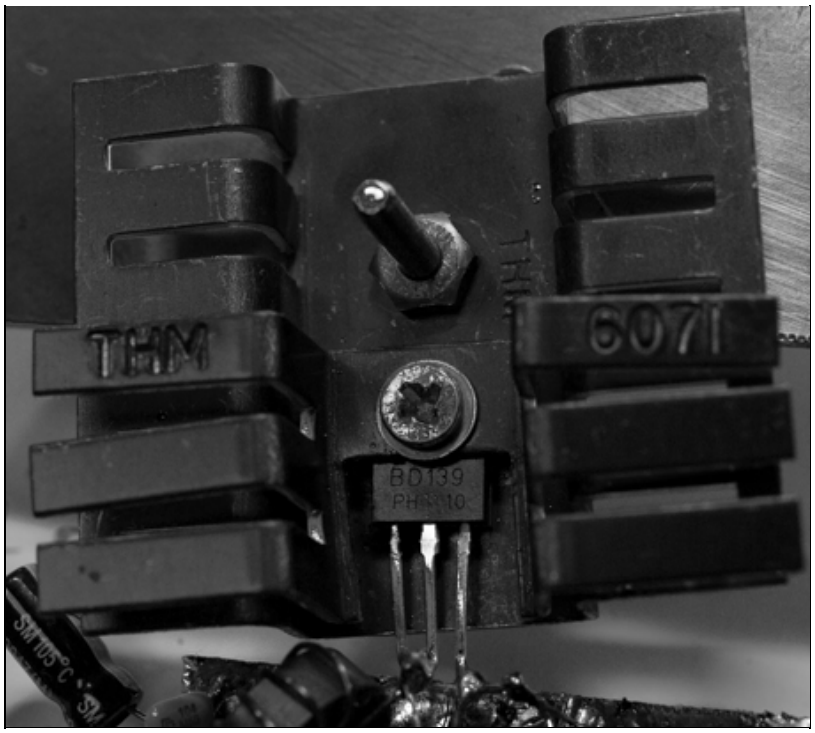
Section 6: Miscellaneous Figures and Photos

8047
C3039
M0H4

2N3553
K0350
2N3553
K0350

2I 06000
WC9838


THM
6071
BD139

RF — Test and Measurement

HF - Embarcadero

Respository web page for HF experiments performed in Fall 2013 and Winter 2014.

1. IMD Oscillator Quest
2. Almost Popcorn Superhet — an older, variant and incomplete HF project added for sake of interest
3. VXO — VCXO Notes
4. Funster Line — A Simple 40M Band QRP Trans-receiver.
5. Miscellaneous Photos and Figures

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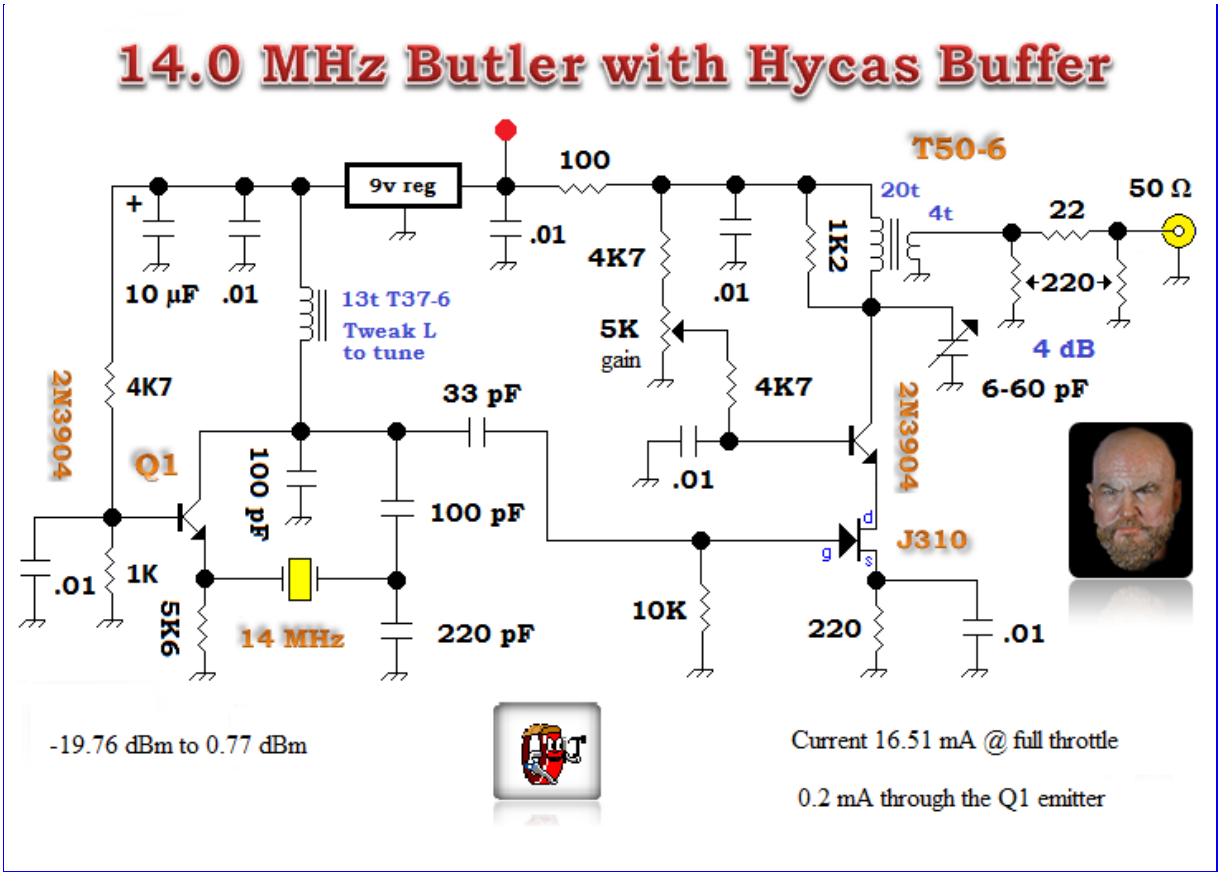
1. IMD Oscillator Quest

To measure IMD and calculate IP3 in receivers and components, we need 2 oscillators that typically lie 20 KHz apart, although some apply other frequency spacings. We join these oscillators with a 6 dB (or 3 dB) hybrid combiner and often follow the combiner with an RF attenuator and/or low-pass filter before connecting to a DUT and detector. I sought an oscillator possessing low phase noise, low distortion, strong return loss + high reverse isolation and documented my experiments to fuel your own. Nestled half-way within the HF band, 14 MHz is a popular IMD oscillator frequency choice. I chose 14.0 MHz and 14.020 MHz for classic 20 KHz spacing.

Wes, W7ZOI well covers intercept measurements in EMRFD and in part, popularized receiver DR measurements with his QST article entitled Defining and Measuring Receiver Dynamic Range in QST for July 1975. Click for the article header only. After publishing his article, receiver DR testing by the major Ham transceiver manufacturers became standard.

If you own 2 crystals spaced 20 KHz apart, you're set. A Butler oscillator set up for vibration at the xtal fundamental frequency works well. See EMRFD Figure 7.32 for a great example — A Butler carefully biased to stand reasonable current for low phase noise with a low impedance in the Colpitt's tank. Wes placed a 100 Ω resistor on the link between base terminals of the differential pair. The low impedance and the bias levelling provides current limiting and prevents the oscillator BJT from going into saturation.



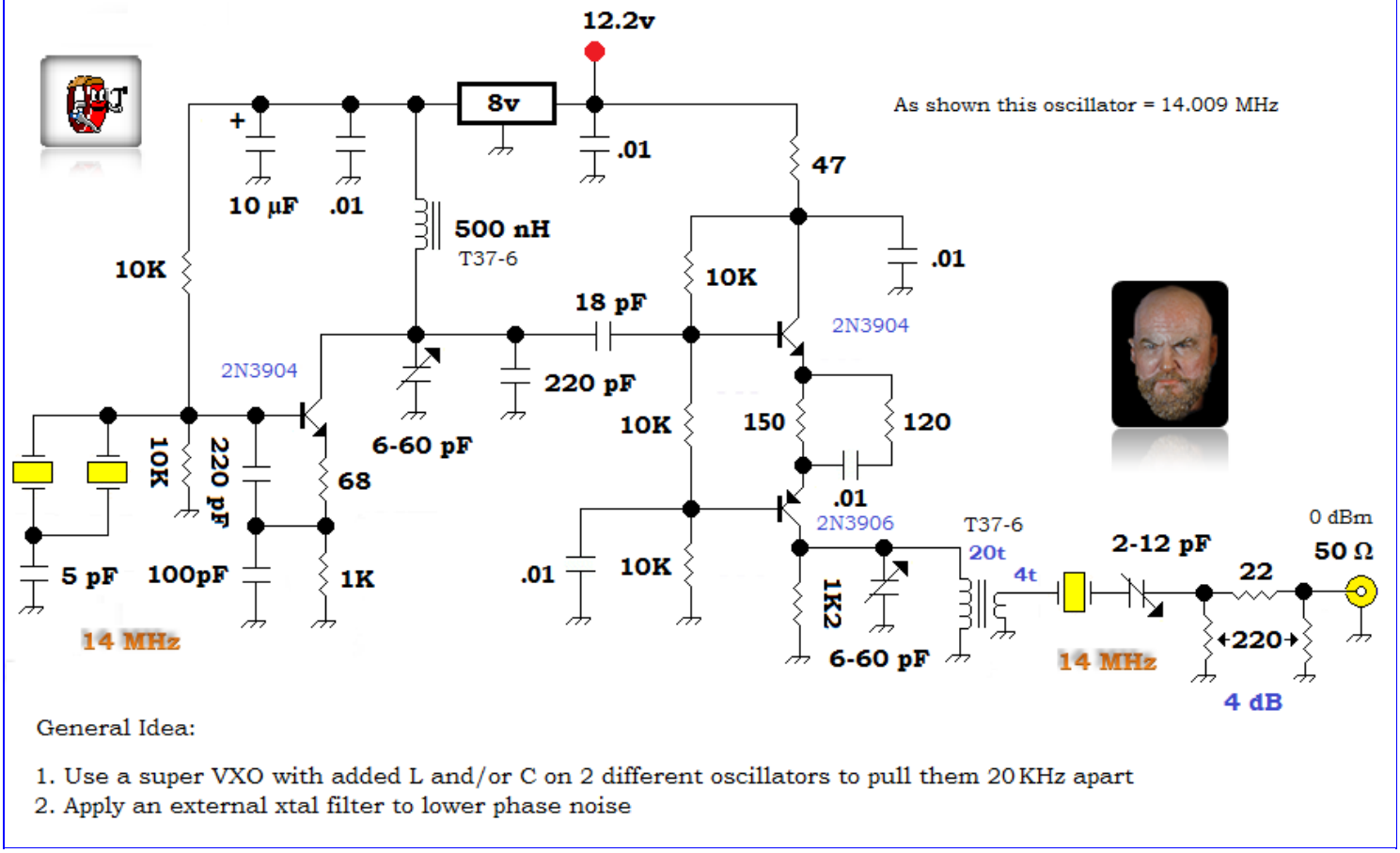
Above — A fundamental frequency Butler with hybrid-cascode (hycas) buffer I built for 14.0 MHz some years ago. Rather than tuning the oscillator tank with the usual low-Q trimmer capacitor, I designed for a fixed C of ~100 pF and thumb and finger tuned the tank to preserve the Q.

Alternately, a standard Colpitt's crystal oscillator with the crystal connected to ground via a shunt capacitor of >= 200 pF provides a great way to get a low phase noise, low distortion oscillator. Take the oscillator output from the junction of the xtal and shunt capacitor and ensure you lightly load this node. See EMRFD Figure 4.24 or my Low Noise Crystal Oscillator web page for examples.

A problem arose. I couldn't find 2 crystals 20 KHz apart, so running 2 fundamental Butler or Colpitts oscillators for IMD measurement seemed out of reach. A web search yielded 1 solution; pull the xtal oscillators to a desired spacing around 14 MHz and crystal filter the output to drop phase noise.

My next experiments mirrored the work of Christian, DL9NL on this web site. It appears he based his design on that of Stuart, KI6QP who authored A Precision Two-Tone RF Generator for IMD Measurements for QEX for April 1995. Stuart used crystals cut 20 KHz apart to make 2 Colpitt's oscillators running moderate current with output filtering by 1 xtal filter tuned to the oscillator frequency to reduce phase noise.

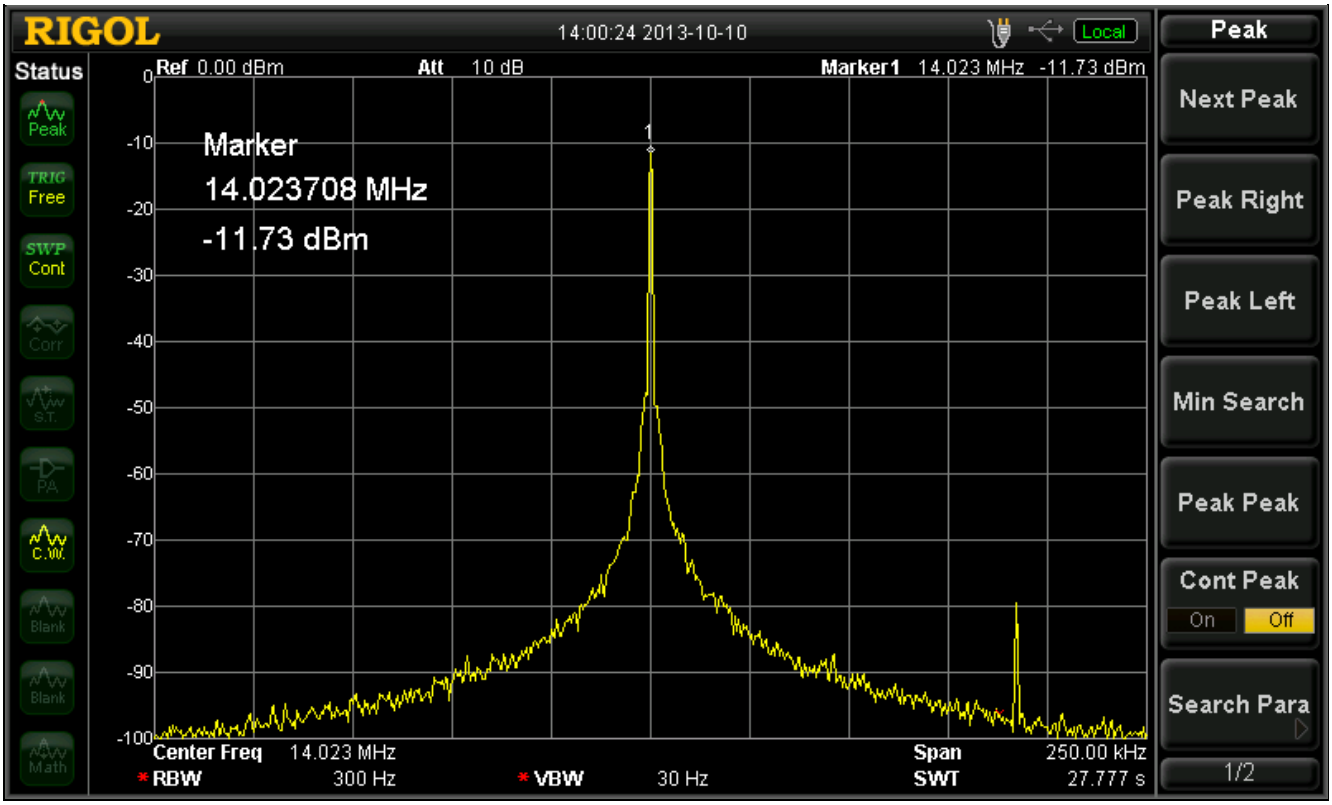
In his project, Christian, DL9NL wired up two 14.318 MHz microprocessor filters in a super VXO topology to cleverly move the xtals 5 or 20 KHz apart. He also output filtered his 2 oscillators with a xtal filter.



Above — 1 of my rudimentary experiments involving a super VXO plus output xtal filter. Not a great design. In this experiment, the 2 crystals were connected to ground via a 5 pF capacitor to yield 14.009 MHz. I learned that xtal frequency and xtal parameters determined the available frequency spread and output filter function.

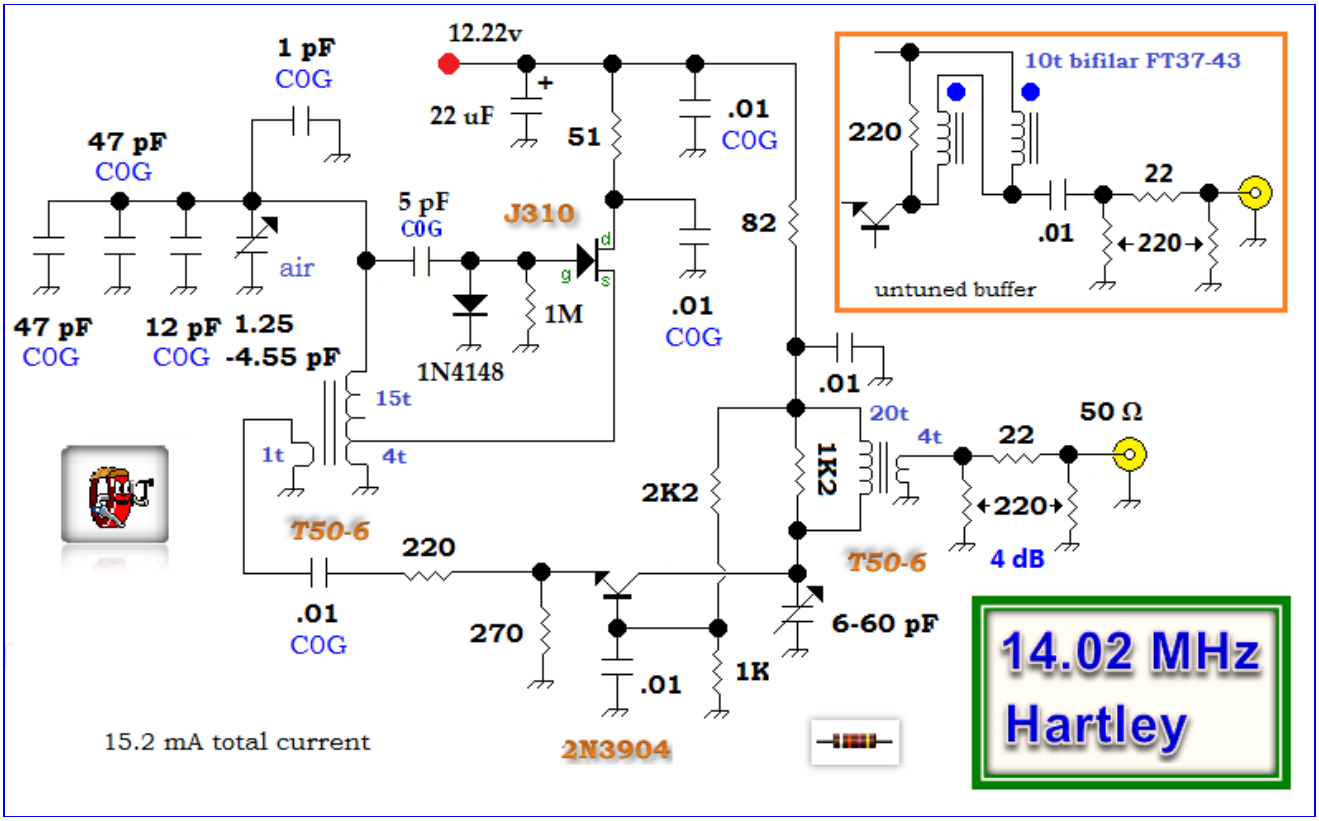
After thought and the above bench work, and also because I need to make oscillators for IMD measurements at 50-100 MHz, I decided to go with an LC oscillator. They work fine for IMD measures with proper design and construction. For example, the signal generator shown in EMRFD Fig 7.27 sports a phase noise approaching -140 dBc/Hz @ a 10 kHz spacing; getting awfully close to the cherished vintage HP-8640B signal generator.

With LC oscillators, I learned that a metal enclosure with a feedthrough capacitor for the DC line proved important.Strong reverse isolation also ranks 'very important'. A case in point follows:



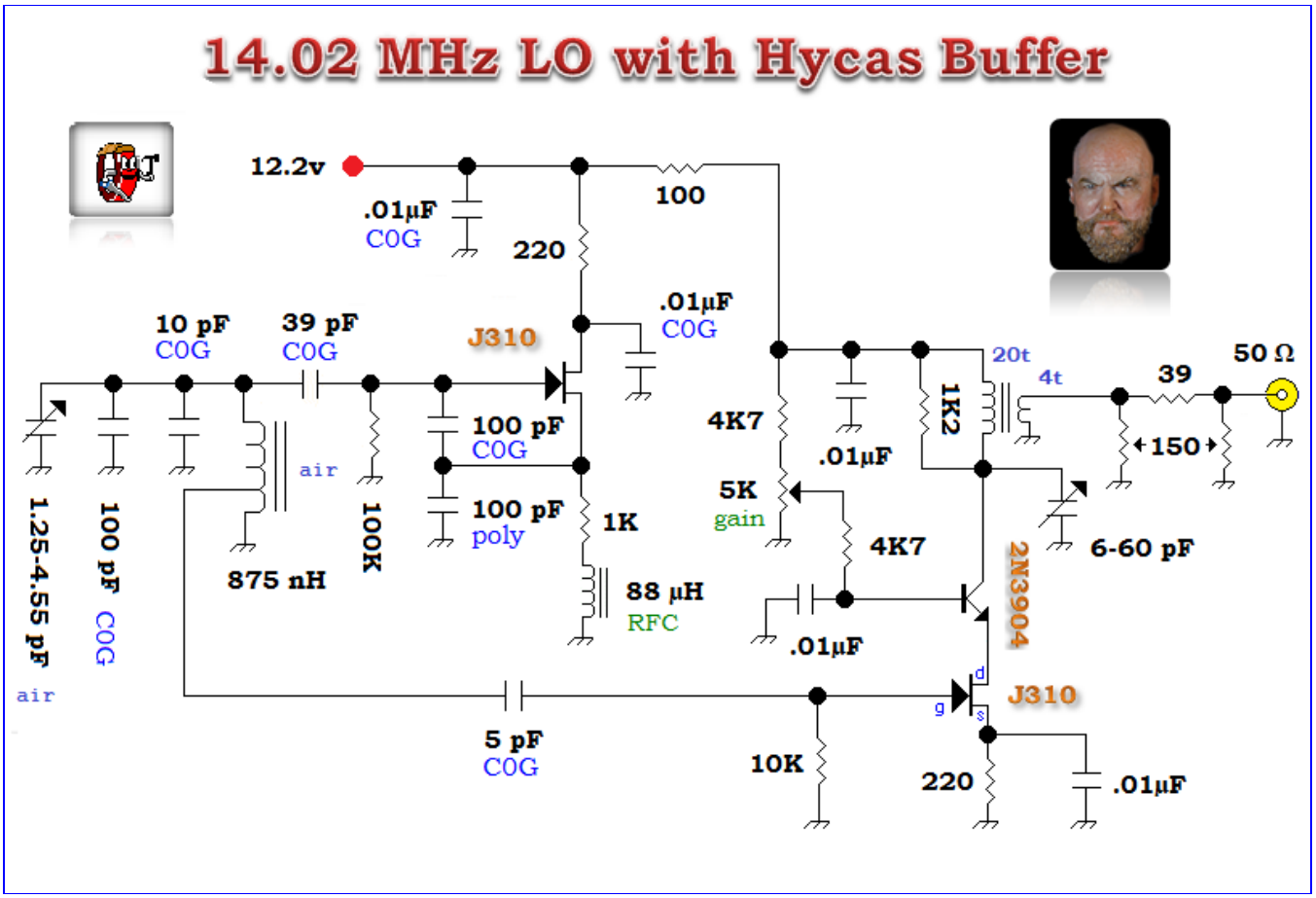
Above — A close-in spectrum analysis of 1 of my 14.02 MHz LC oscillators. Span = 250 KHz; RBW = 300 Hz and my video bandwidth = 30 Hz to "cut the grass". You can see a spur about 92 KHz right of center; a signal from my SA or environment is getting back into my VCO and low-level modulating it. Yikes! Another RF-tight LC oscillator with a common-base isolation amp showed no such spur.

Close in spectrum analysis may allow you to measure phase noise with your spectrum analyzer, however the wide-range, first oscillator of a Rigol spectrum analyzer operates in the range of 2- 4 GHz and is varactor tuned; as such, the best phase noise we can measure with it = -88 dBc/Hz @ 10 KHz spacing.



Above — 1 of my experiments: a Hartley with common-base buffer amp right out of EMRFD Figure 7.27 sans the final amp. I made 2 versions: 1 with a tuned buffer amp and another with a wideband output (see orange inset). The output power of the tuned amp and wideband amp = -6.7 dBm and -11.7 dBm respectively.

You may want to add a final amplifer — I've got a couple of amplifer modules if I need to get to 0 dBm or above. Prior to tweaking the lid-on frequency to 14.02 MHz, I measured the 2nd harmonic at 55 -dBC in the tuned output version— yay! Building with solid temperature drift techniques, the frequency drift in my counter went just a few hertz *up* over time. The Figure 7.27 oscillator remains a go-to circuit for us builders.

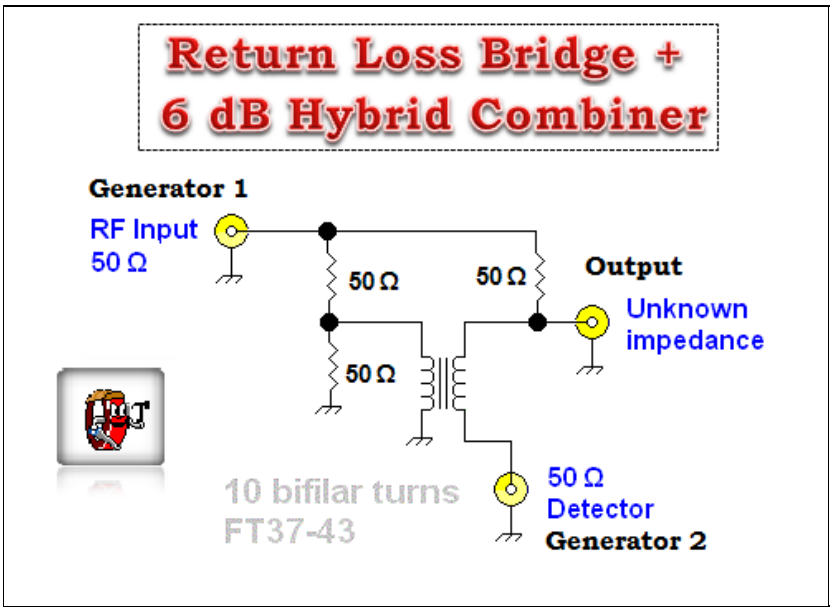


Above — A Colpitts with hycas buffer allowing variable power output. During measurement of IMD, the 2 oscillators need to exert the same amplitude, so only 1 needs to have variable output power. A wideband bench generator plus a 1 frequency oscillator like this might be all you need.

Above — A photo of the completed Colpitts LC oscillator. Running a phono jack for the DC input was a mistake. A feedthrough capacitor = the best choice. The epoxy glued air wound coil and air trimmer cap plus the regular temperature stability techniques and a 100 pF polystyrene

capacitor yielded a temperature drift under 40 Hertz per hour. 5 bolts anchor the copper clad board.

I connected the 5 pF coupling cap to the lowest turn on the air coil. Mounting an air coil (solenoid) in smokestack fashion over a copper board probably upsets its Qu, but it worked okay.



Above — Our old friend the Return Loss Bridge is also known as a 6 dB hybrid combiner and 1 serves double-duty on your QRP workbench as an isolator and a return loss bridge.

Phase Noise and Commercial Signal Generators

The worst of today's ham transceivers transmit a CW carrier with noise that is down by 117 dBc/Hz @ 10 kHz spacing. This is bad enough that if you had a Ham operating this rig 0.8 km away you would hear this noise, even with a perfect receiver. In a typical oscillator, the phase noise @ 100 kHz spacing lies ~20 dB lower than @ 10 kHz out to a noise floor that is around -140 dBm/Hz or better. The Elecraft K3 impresses us with phase noise at -138 dBc/Hz @ 10 KHz [Sherwood].

We either build, borrow or buy our signal generators. Second hand commercial products range from good old HP/Agilent gear or LC boxes once used by TV/FM receiver repairmen. The TV service equipment covers ~300 kHz to 150 MHz and tends to be inaccurate with a limited output amplitude range, while the vintage HP stuff often is heavy, decaying and crowds your bench.

Most of hobbyists can't afford nor justify spending $8-18k on boxes that go down to -127 dBm amplitude and -146 dBc phase noise at 1 GHz @ 20 kHz offset [Agilent N5181B MXG X-Series RF Analog Signal Generator]. These generators feature components like a YIG oscillator phase locked to an ovenized crystal reference. Professional, high-end gear.

Another choice = the Arbitrary/Function Generator or ARB that sell for a fraction of the cost of the aforementioned boxes. A few friends own them, in particular, products made by Rigol and GW Instek. These DDS generators range in frequency and price: higher price generally = better phase noise.
For example, the popular Rigol DG1022A phase noise = -108 dBc/Hz @ 10 KHz offset and the DG4162 = ≤-115 dBc/Hz @ 10 KHz offset; not good. Often ARBs will feature a host of modulation functions, sweeps, frequency counters and the 2 output channel models may offer I/Q outputs and the ability to perform 2-tone IMD tests. Search around the T Equipment website to read some equipment datasheets

My question to 2 informed EEs — Is the phase noise of a -115 dBc/Hz ARB *good enough* to reliably measure IMD? Their answer: yes, if you have a high end, modern spectrum analyzer with a minimum RBW of 100 Hz, deft analyzer skill and understand the measurement. Often us hobbyists have to make due with whatever test equipment we can build or buy.

You might just build a low noise oscillator with a crystal at 14.0 MHz and provide the other tone with your bench signal generator or some other combination of 2 devices. This is test equipment junkie fodder — band-of-brother's stuff. Thanks for reading.

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2. Almost Popcorn Superhet

I published this rogue web page with hope you might find it interesting. It houses notes and photos for an incomplete update to my 1998 Popcorn 7 MHz, CW, superhet receiver.

Click for the Almost Superhet web page.

Click for the Almost Superhet Supplement page

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3. VXO — VCXO Notes

VXOs and VCXOs provide a useful main or reference oscillator for HF and when multiplied, in VHF-UHF designs. Ignorant of such circuits, I executed some crude experiments to discover how they work and find ways to improve my designs. Increasingly clinical on my bench, I'm losing

patience with the radio mystique — although the limits of my lab equipment, brain power and bench skills humble and tumble me into deeper understanding why educated people might submit to electronics folklore and go in blind faith.

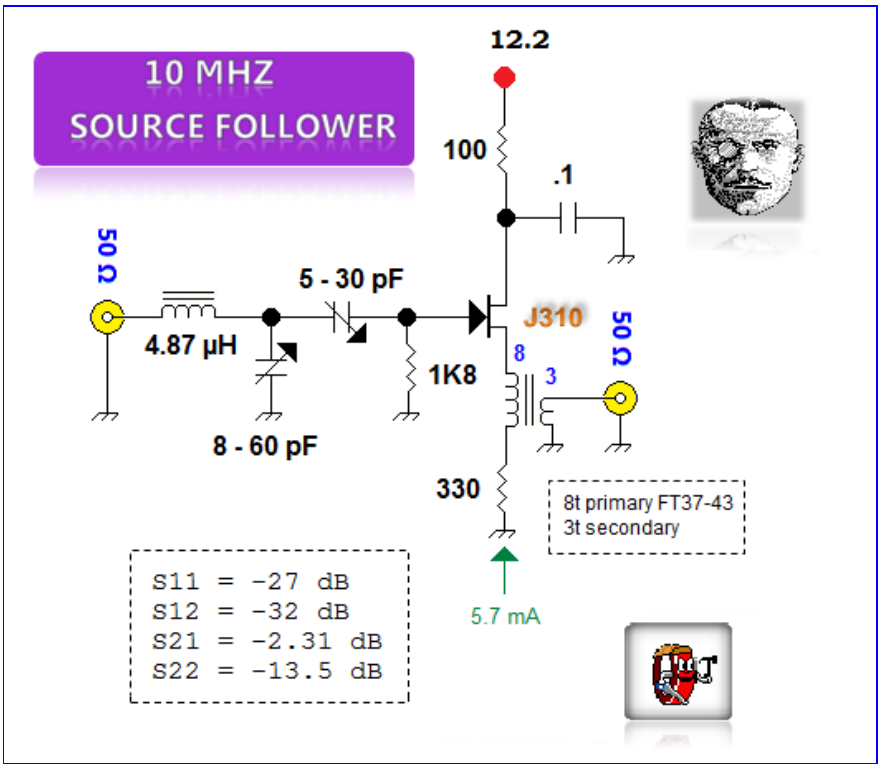
I'm not sure my bench work accomplished much, but if I can add 1 factoid to the RF enthusiast VXO knowledge base, then, I'll feel happy. I'm officially adopting **System D**; the slang term that originated in the French-speaking African continent and Caribbean Plate. In Down and Out in Paris and London, George Orwell called these System D practitioners "débrouillards" — low-level kitchen help, who achieve culinary success in the worst of circumstances. Television hero MacGyver practiced System D.

System D in the context of the QRP Homebuilder describes an effective, practical and resourceful builder who skillfully adapts to any situation and *gets it done*. System D builders don't just kludge together crap — these science-informed débrouillards innovate and shine. Sounds hopeful doesn't it?

Nearly all my bench adventures begin with a web search and the key word *VCXO* yielded many espléndido files + presentations by people and companies who passionately share their knowlege and experiences. Thanks to all!

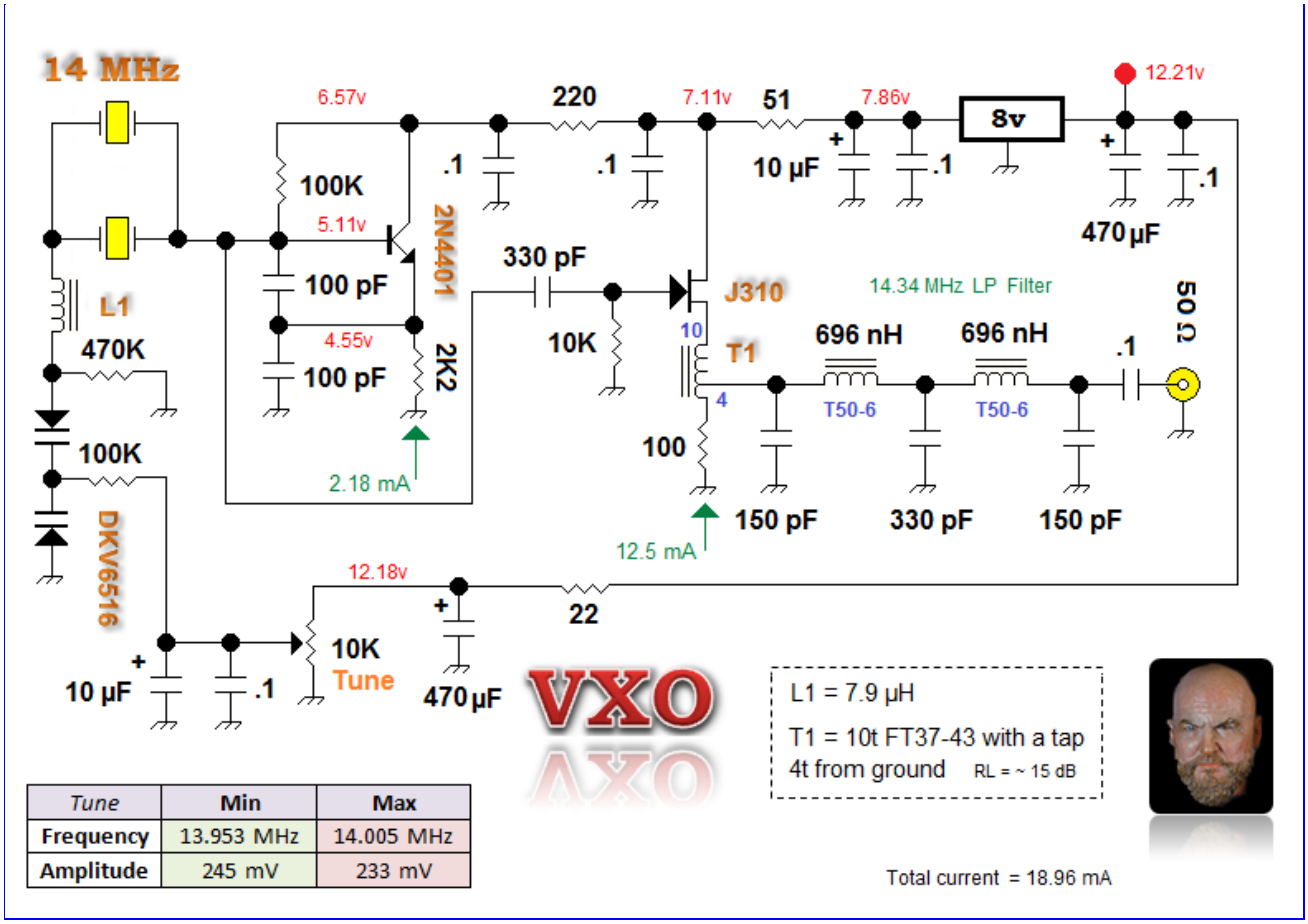
Experiments

It started with an email from Paol — what's the S12 of the source follower in his 14 MHz VXO buffer amplifier? I built 1 for 10 MHz and tested it since I didn't know.



Above — Device Under Test. Paol copied a source follower he found on a web site (complete with an über kitsch 1 mH choke on the source). I tweaked the current and replaced that big, old MF choke with a tapped FT37-43 giving an output return loss of ~ 13.5 dB. Although mediocre, it's probably better than the match provided by the series resistor we typically stick between the JFET source and its 50 Ω load. Further, for my S12 measurement, I matched the input with a L-C-C network to get a strong return loss @ 27 dB. Measured with a spectrum analyzer, S12 = -32 dB; paltry reverse isolation. The matching networks incurred some losses but raised measurement fidelity.

-32 dB — not great, but okay if followed by another amp. For a diode ring mixer, I might couple it to a common base amp, or eliminate the source follower altogether and replace it with a hycas buffer amp. Paol drove a NE612 transmit mixer with his VXO + a source follower and it worked okay in his bare-bones TX. He AC-coupled a 51 Ω resistor to ground on pin 6 of the NE612 that well terminated the LP filter. Intrigued, I then helped him with his VXO.

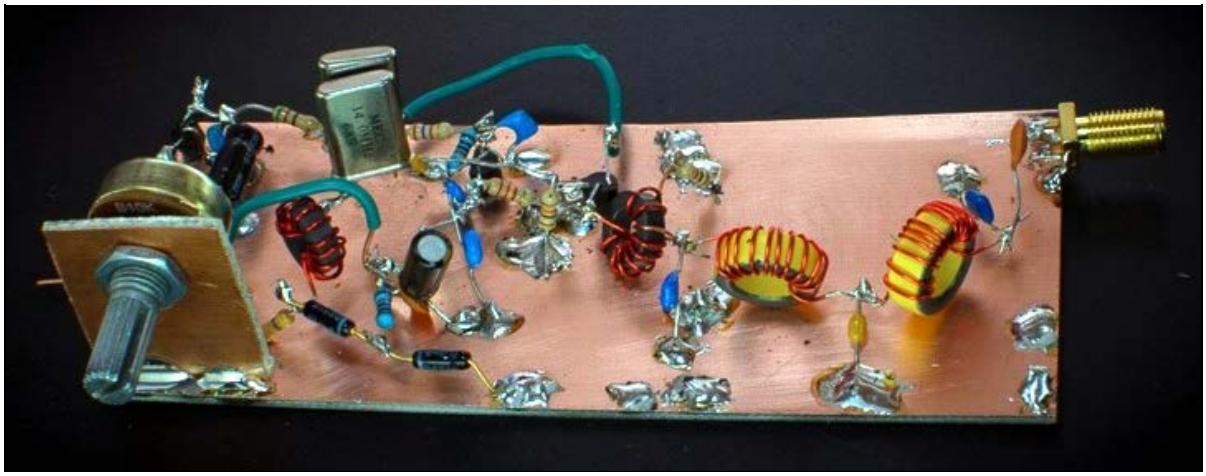


Tune	Min	Max
Frequency	13.953 MHz	14.005 MHz
Amplitude	245 mV	233 mV

Above — My first VXO experiment with varactors instead of an air variable cap that technically creates a voltage-controlled crystal oscillator (VCXO). The tabled values were measured with L1 wound on a FT50-61 ferrite toroid.

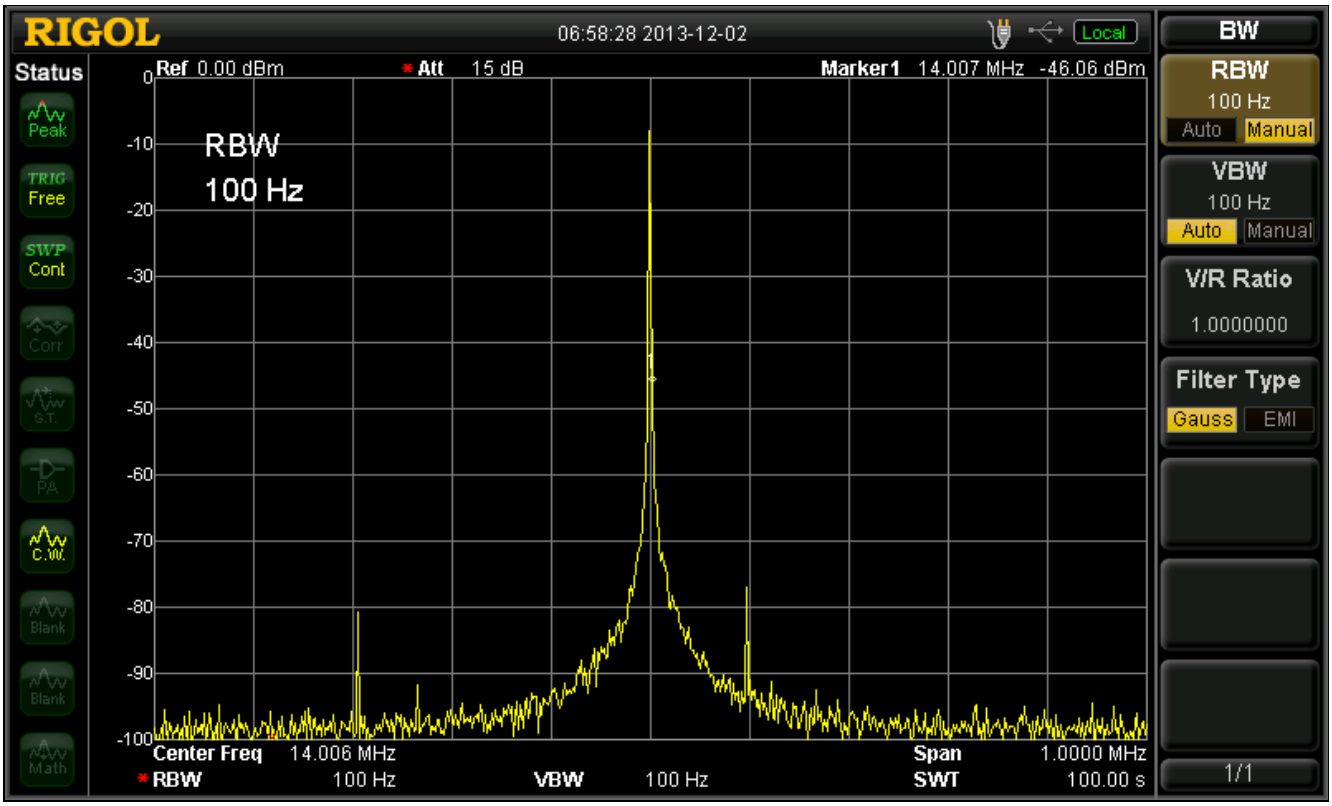
Paol had varactors but no air-variable capacitors. Both of us owned some Q = 200, DKV6516 silicon varactor diodes that with a reverse DC voltage of 2 to 10 volts give a nice tuning range of 18 to 140 pF for a single diode. We cheated and ran the reverse DC voltage from 0 to 12.2 volts, although later increased the 22 Ω R to limit the peak tuning voltage to 10v. Paol also put a fixed resistor on the cold end of his 10K tuning pot to keep a minimum 2 VDC on the varactors: I didn't bother, since my "this is only a test" breadboard will be dismantled and junked.

Above — DKV6516

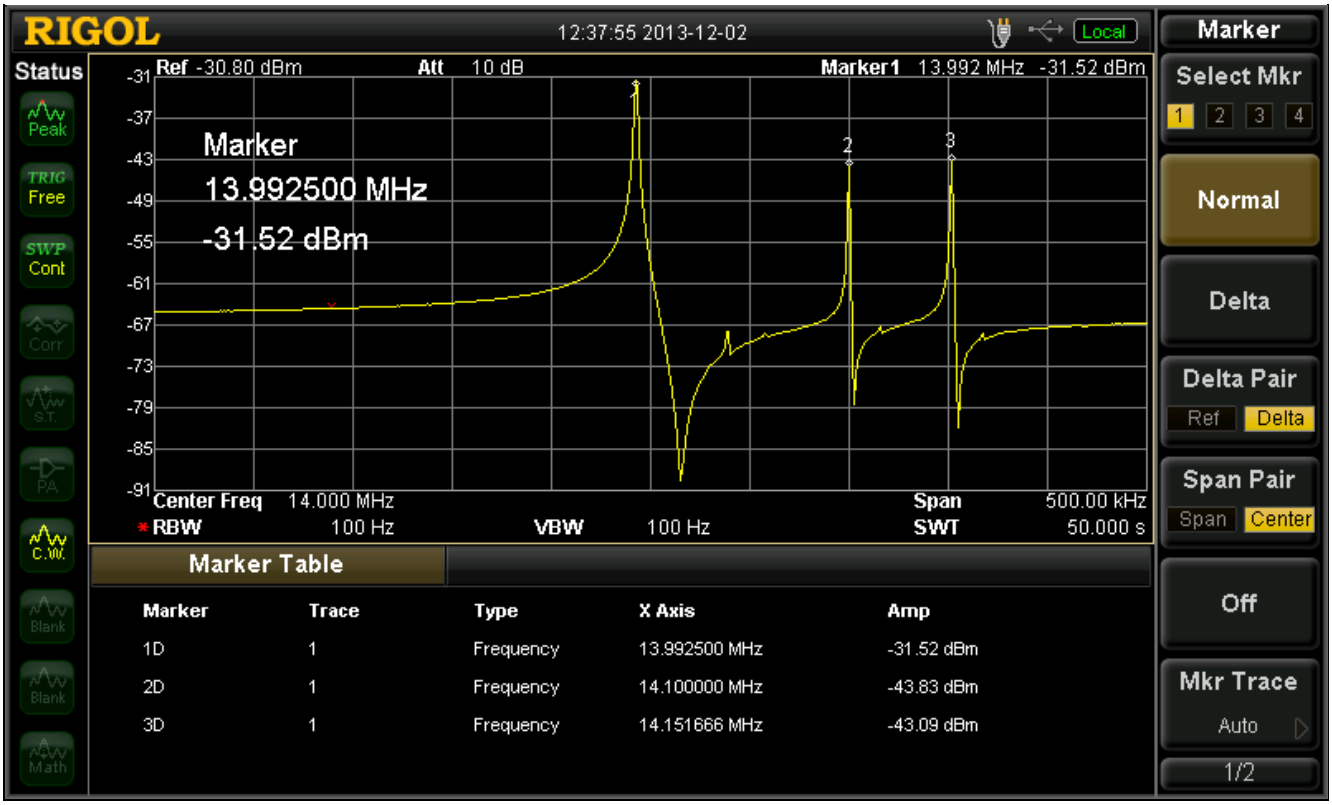


Above — VCXO breadboard. Paol and I each made a VCXO and he with 14.060 MHz crystals. Lacking these, I just used 2 junk box 14 MHz crystals with a measured Qu of ~80K. The tapped FT37-43 gave an output return loss of ~ 15 dB thus providing a reasonable match into a simple 50 Ω low-pass filter. Click for a sweep showing the 2nd harmonic suppression.

My initial build didn't contain the two noise filtering 470 µF caps on the 12 volt supply: I originally soldered a solitary 22 µF cap cap close to the 8 volt voltage regulator input. My 12.2 volts comes from a regulated DC supply. More on filtering noise later.

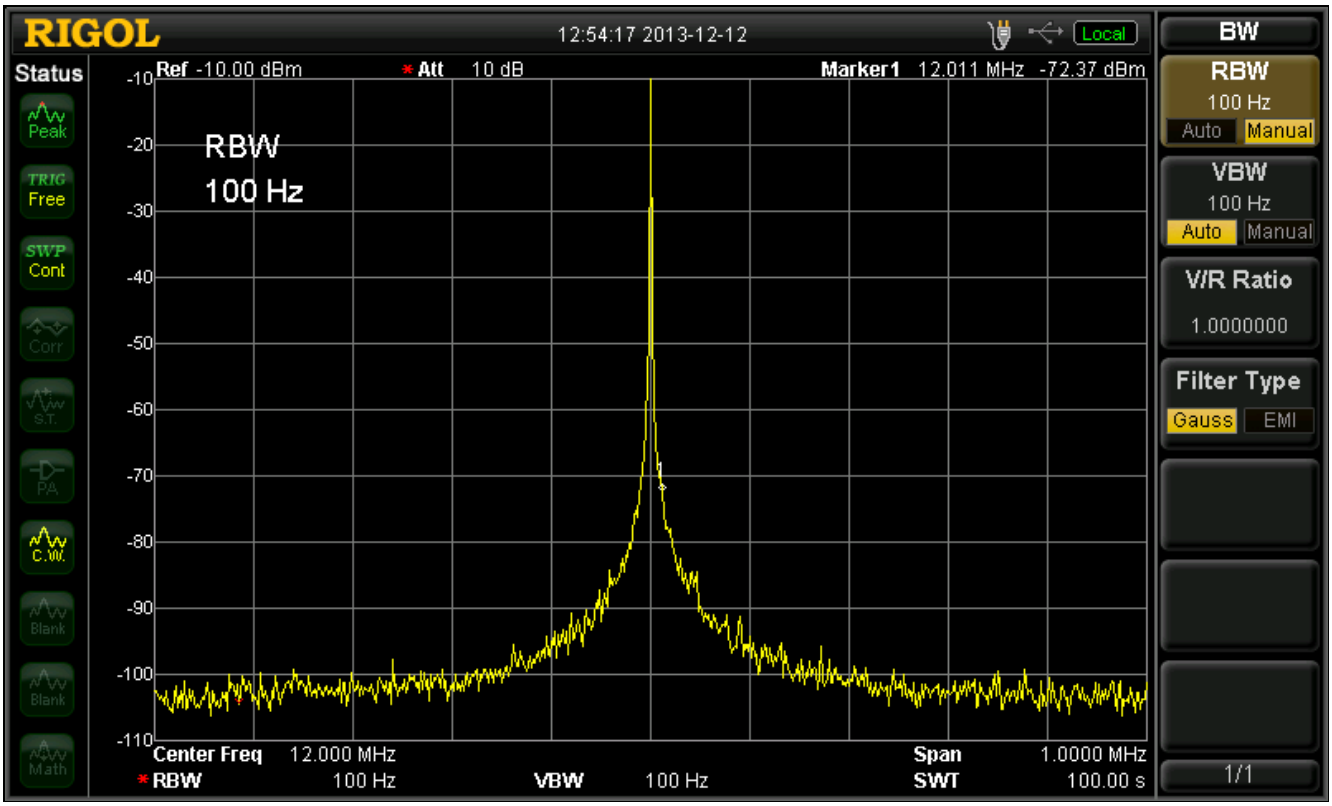


Above — My VCXO spectrogram showed 2 big spurs: 300 KHz down and 100 KHz up from center frequency (CF). I exchanged the varactors with an elegant, Q >= 1000 air variable, plus changed to a higher Q inductor, but could not get rid of these spikes. I assumed these spurs came from the crystals, or the BJT. Then, too, might they arise from my spectrum analyzer?

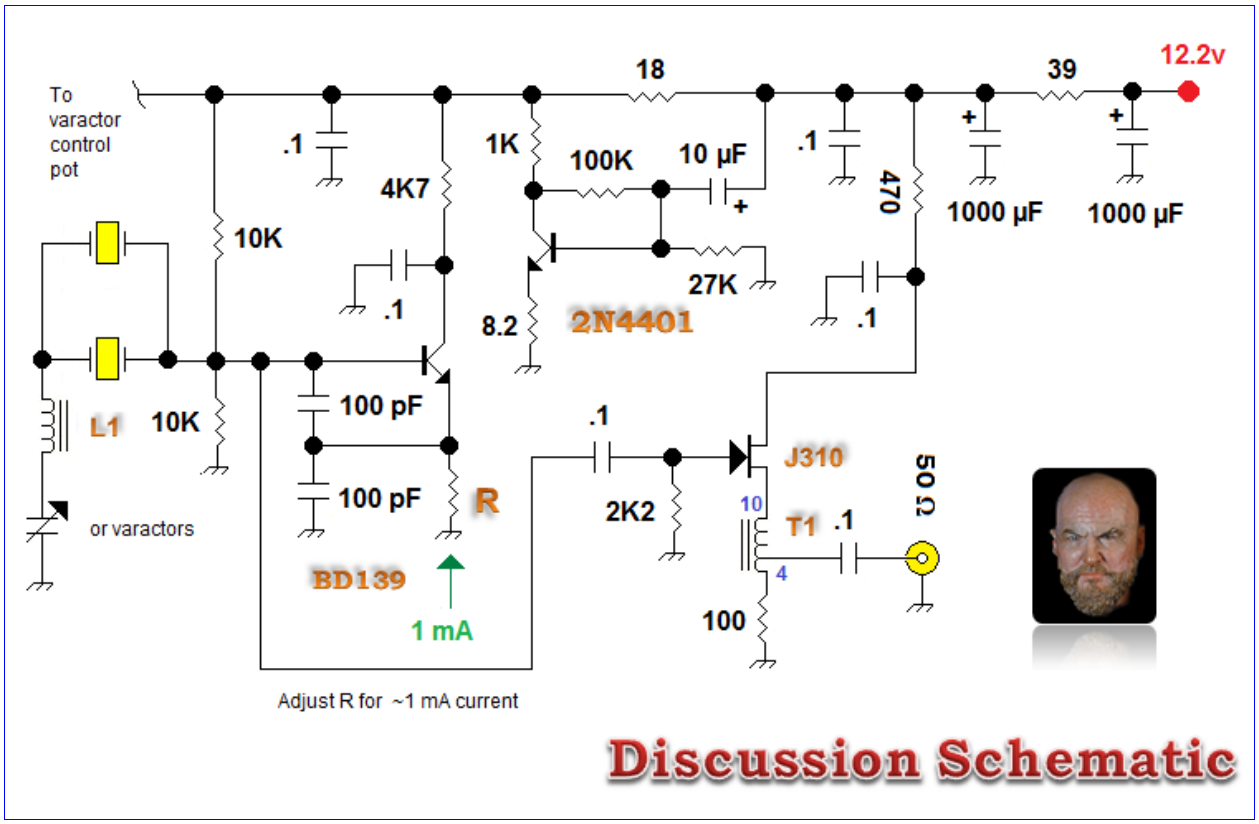


Above — 14 MHz crystal sweep spectogram. To find out why these spurs occurred, I swept 1 crystal at a time from my batch and saw spurs on all of them. However, after some reading, I discovered that the mere presence of a crystal spurious response doesn't signal causality since this is almost a normal happening. Wes, W7ZOI explains this on this web page. Crystals usually have a spurious response and from Wes, the spur is more often than not about 80 or 100 kHz above the main signal. This happens with fairly good crystals, less so with crappy, cheap crystals and even less often with good quality crystals. Since many of us buy computer-grade crystals from Mouser-Key, or online auction sites, who really knows what we're getting? 1 day I hope to test a Q = 1 million crystal — that will be fun.

However, renograde crystals could still be the cause. Sometimes when pulling a crystal(s), we're living on the edge — the main resonance is pulled so far that the circuit balks at the resonant frequency, or lies near the edge of such behavior and spurs start to show up. *Back to System D*. Simple substitution with 2 other crystals from another batch could tell me if these particular crystals were behaving poorly. It turns out, they were.



Above — Spectogram of the VCXO with 12 MHz crystals. I wanted to try other 14 MHz crystals, but didn't own any. My nearest value crystals were 12 MHz jobs with a Qu of 120K. Eh voila! No spurs — yay! Let's talk about noise:



Above — Schematic for sideband noise discussion

System D design poses a real challenge with repect to noise. Measuring noise proves an advanced technique beyond the scope of this popcorn RF web site. We can't measure noise with our spectrum analyzer since the oscillator in the analyzer usually exhibits more noise than the VXO we want to test.

On page 4.12 of EMRFD version 1 is a small photo of 2 spectograms that show a clean crystal trace contrasted with a really bad oscillator trace. Baffled, because I could not replicate this experiment, I emailed Wes and asked him how he did it. Wes wrote we're able to see the noise in those traces because his SA featured a multiple pole, wideband filter with steep skirts that allows us to see the baseline noise close-in to the carrier. In contrast, most modern SAs feature synchronously tuned resonator type, or digital filters that produce a narrow peak with skirts that spread way out. The noise gets swept under the skirts.

By all means, advanced builders may apply notch filter techniques to measure oscillator sideband noise down low, however, a consideration for others might be to apply techniques proven to reduce noise and spurs — and just build the best oscillator possible plus measure what you can.

Some basic techniques to make a low-noise VCXO are well described elsewhere and may include: a high Q resonator, high Q varactors [with low series resistance], low power supply noise [including varactor tuning voltage], a low flicker noise amplifer, avoid overdriving the crystal and try to swing the highest AC voltage in the tank without reaching saturation and/or causing reverse breakdown in the varactor(s). Even the ratio of the Colpitts capacitors can affect phase noise, but without a method to measure phase noise, this gets esoteric. Some of these techniques also apply to VXOs.

Power Supply Noise

Bob, K3NHI and others have measured significant noise from the LM317 or related voltage regulators that we amateurs typically use. Power supply noise can FM our resonator and boost oscillator sideband noise. We may filter power supply noise with *large* value capacitors plus/minus series decoupling. See the work of Leif, SM5BSZ on this web page under the heading "The +12V supply is carefully decoupled like this". In addition, Charles Wenzel, published this web page.

I employed both techniques in my discussion schematic above. I went with the pi filter shown above 1000 µF - 39R - 1000 µF at the voltage input. A inductor might be used instead of the R if the voltage drop is too high for you. The Wenzel circuit fascinates me. The series resistor (18 Ω in my version) samples the current that flows into the oscillator. A voltage is then developed that is applied to the amplifier to generate a voltage that is out of phase with the noise current.

With good DC noise filtering I could measure small improvements in sideband noise in my SA despite only able to measure down to -88 dBc/Hertz at 10 KHz offset. In contrast, you rarely ever see even 1 electrolytic capacitor in the VXO schematics featured on many Ham radio builder sites.

Another trick is to avoid a regulator altogether. Since my DC supply already has a voltage regulator set to 12.2v, placing a second 5 to 8 volt regulator after it might just add another noise source.

Overdriving the Crystal

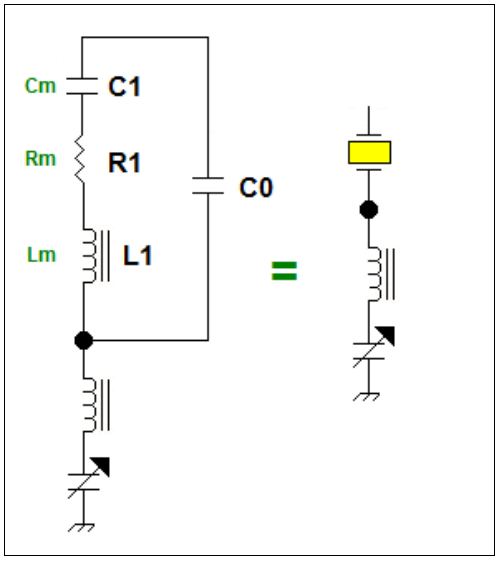
I heed the advice of Konstantinos, SV3ORA. He found that gently driving his crystal generated less noise and spurs compared to overdriving the crystal. Thus I adjusted my emitter resistor to about 1 mA current. For the schematic above, a typical emitter resistor starting value = 1K8. Just measure your emitter current.

I employed a large area, slower BJT, the BD139 to hopefully drop flicker noise. Expensive, super high Q crystals (Q = 500K to 1 million) may be driven with up to several mA of current, however, it's likely not so for the computer-grade xtals we typically use. I can't measure the outcome of these interventions, but System D thinking might just prevail.

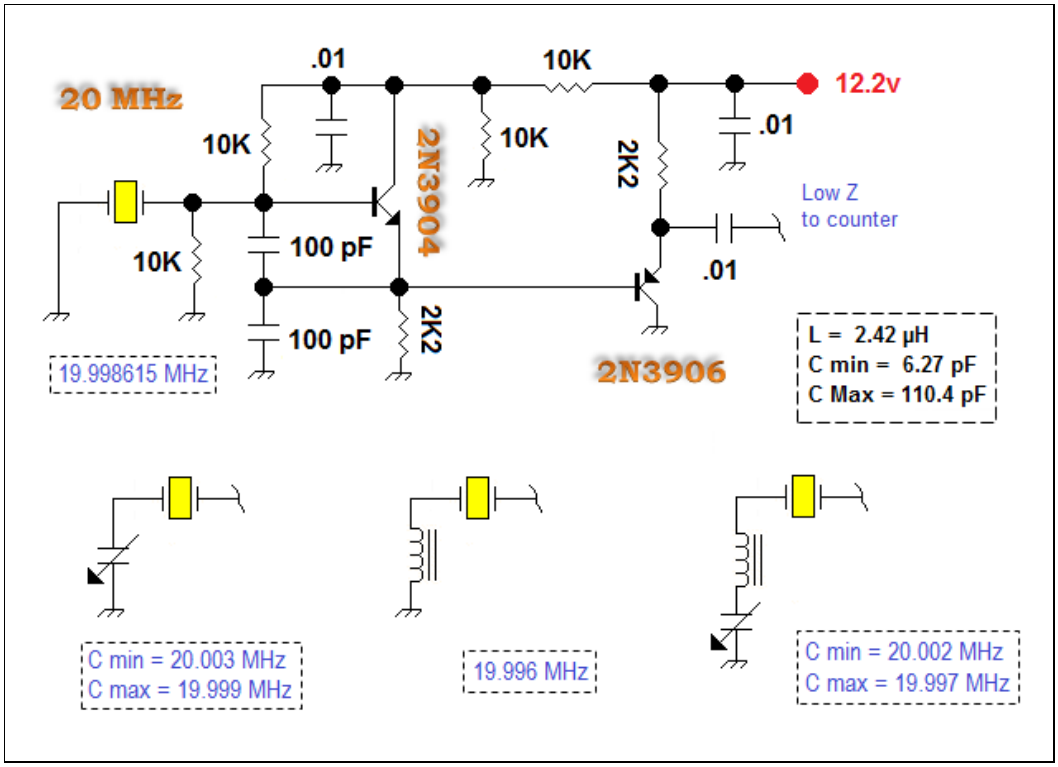
Varactors

Varactors tend to be low Q and their Q deteriorates as we go up in frequency. I've seen very low phase noise VCXOs employing varactors, however, these were engineering marvels. Still, as a VHFer and hopeful UHF enthusiast, this is the direction I wish to go. We ideally need to keep some reverse DC voltage on the varactors at all times and not go to 0 — this also limits tuning range in our simple 12v supply radios. Read your datasheet. Air variable capacitors (a VXO) might be a better solution for your application.

How Does a VXO work?



Above — The crystal equivalent electrical circuit has 2 arms: The impedance of the motional arm Zm lies in parallel with Zo, the impedance resulting from the parallel capacitance C0 (formed by the 2 crystal leads.) We measure and/or calculate these parameters when designing our crystal filters. In the above schematic I added the external L and C used to move the crystal's series resonant frequency around in our VXO, VCXO, or Super VXO circuits.



Above — I made some simple experiments to explain why we place both a L and C in series with the crystal and how it changes the crystal

frequency in our oscillator. From these experiments, some theory from EMRFD and wisdom from Ken Kuhn, I made the following simple conclusions:

Electrically a crystal looks like a large inductance in series with a small capacitance and in parallel with C0. An external inductor adds to the crystal's series inductance and *lowers* the resonant frequency.

The external variable capacitor adds in series with the crystal internal capacitance to lower the net capacitance and *raise* the resonant frequency.

The purpose of adding an external inductor is so that we can tune both lower and higher than the natural crystal frequency. In contrast, if we only apply an inductor, the oscillator frequency would always be lower and if we just apply a capacitor, the oscillator frequency would always be higher.

Since the external inductance has much lower Q than the effective internal inductance, it adds to noise and signal losses. The same goes for the external C.

As we raise the inductance ever higher to boost the delta F, at some point, the parallel C of the crystal resonates with this "too high" L and the series resonant frequency has nothing to do with the series resonance of the crystal. This kills the normally sublime frequency stability of a crystal-based oscillator.

Q of the VXO/VCXO Inductor

I experimented to learn that the Q of that L doesn't need to be that high to allow the frequency to be moved up or down by an air variable cap or varactor.

For example, if I wound a coil on a T68-2 and shunted it with a resistor no less than 1K8 — it worked fine (sometimes the frequency shift would even improve 2-3 KHz for a given L). A #61 ferrite toroid wound inductor also worked well and requires much less turns allowing thicker wire for mechanical stability. Also, the lower number of turns helps avoid significant SRFs. Temperature stability remains a question with the this and all ferrites.

An FT37-43 or FT50-43 ferrite inductor worked poorly – I measured high signal losses plus the maximum frequency swing dropped when compared to the same inductance value wound on a T-68-2, or FT-50-61. Avoid #43 ferrites in any resonator circuit unless you're purposely building a noisy oscillator for learning purposes.

Some guys use those little molded chokes with a Q of ~ 50 for the L and they also work well. Predictably, as I lowered the Q of the external L, or went from an air variable C to a varactor, signal losses increased resulting in lower signal output amplitude.

Decreasing resonator Q increases phase noise. For a popcorn radio, I'm not sure the change in phase noise would be a deal breaker in some homebrew projects, but as possible, I think it's wise to employ high Q resonator components: crystal, inductor and the variable capacitance. Thus I'll wind my VXO/VCXO inductors on powdered iron toroids hereafter.

Since a quartz crystal resonator is driven by the piezoelectric effect, fundamental plus overtone frequency modes (3/5/ 7/11) may be derived. I unbypassed the collector and placed a collector choke on it to examine the output with my 'scope. The waveform rung with many harmonics. Lots to learn.

Go débrouillards!

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4. Funster Line — A Simple 40M Band QRP Trans-receiver

Click for the Funster transmittter notes.

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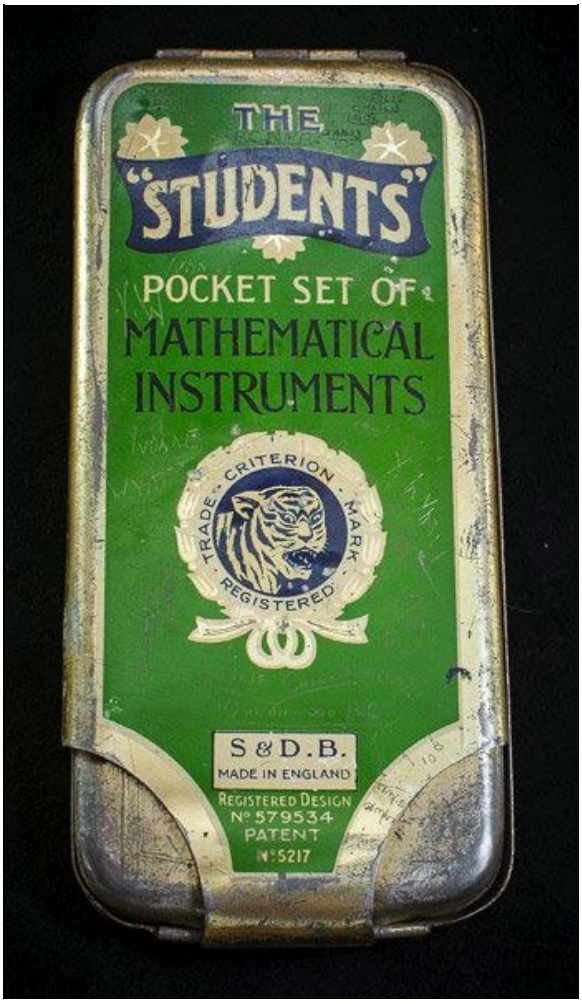
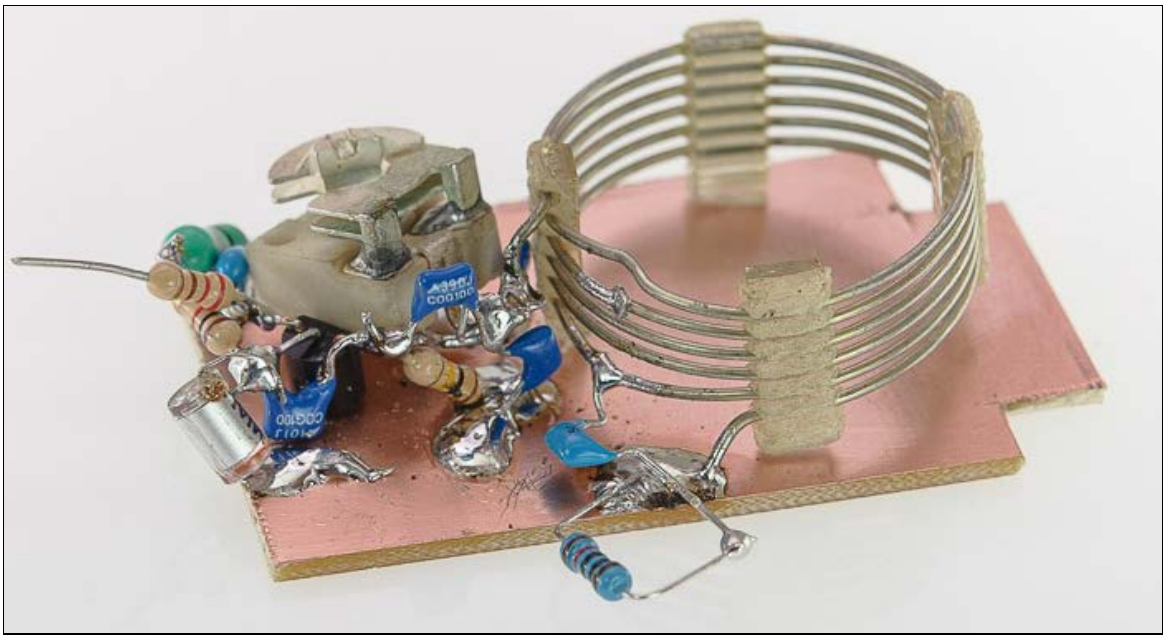
5. Miscellaneous Photographs and Figures

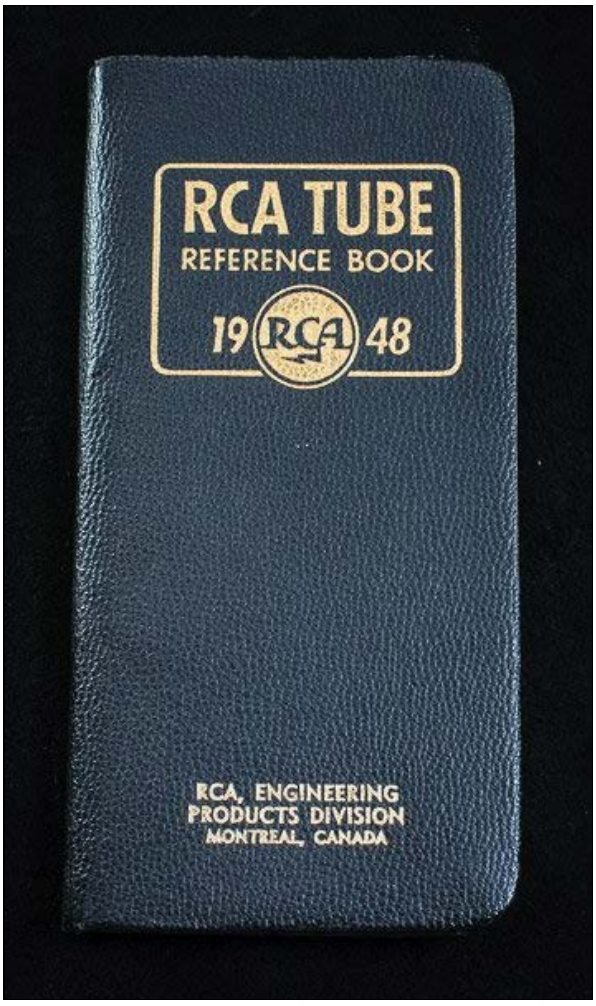


Above — My handed-down geometry set — purchased in 1938. Click for larger picture.



Above — My prototype Colpitts LC oscillator with only a 10K resistor as the load.

Above — Hi Q glass encased crystals. Click for another photo.



April 2014: New player on the scene — Bar Bar Kitty — chief science officer and QC consultant.

April 2014: Other new pet: Munchkin. Clear-spoken and direct — pet me, or feed me... and oh yah, Q2 is oscillating.

MP143
14M3181B
GTSC0313

Caitlyn 310 — UHF Beginnings

Respository web page for beginner UHF experiments in 2014-2015.

1. References
2. Transistors, RFCs, Capacitors and a 50 Ω Detector
3. Return Loss Bridge Experiments
4. VCO Experiments
5. Frequency Counter Kit

1. References

After a modest quest, I found an excellent microwave primer called Microwaves Made Simple. Principles and Applications by W. Stephen Cheung and Fred H. Levien. Published in 1985, the authors brilliantly teach the basics in plain language (no small task). My copy was printed in December 2013 by Artech House who print single copies of many classic texts on demand. Click for a photo of my book.

The ARRL published 2 UHF compendium books entitled: The ARRL UHF / Microwave Experimenters Manual [1990] and The ARRL UHF / Microwave Projects Manual [1994]. While out of print, used copies seem prevalent and might serve as good resources for builders who hunger for circuits and Ham-related UHF circuit building information. Click and click for photos.

Instrument companies provide continuing education and materials to educate & support their customers. With luck you'll find helpful free application notes and other tools. Two recent RF library additions include an Agilent CD containing all their current application notes and webcasts [my image is truncated and the CD includes applying SA's, 'scopes and signal generators], and this old Spectrum Analyzer Fundamentals booklet by Tek.

2. Transistors, RFCs, Capacitors and a 50 Ω Detector

I want to grasp circuits from 300 to 1296 MHz. This meant acquiring some UHF parts, making some bench tools and buying a 50 Ω detector to sweep and/or measure power in my UHF stages. Click for a snapshot of some of my parts cabinets.

BJTs

I'll employ the BFR93A, 2SC3356, and 2SC3583 as my mainstay NPN bipolars; at least to start with. Numerous transistors will work — price, availability, NF and shipping costs informed my choice. I'd also feel happy with MRF901, BFR92 and BFR96 NPNs and it's always fun reading datasheets. My PNP choice = BFT93 by NXP. Later on, I'll add in discrete GaASFET and other non-BJT species. Only 2 MMICs lie in the parts cabinet: MAR-3 and MGA-61563.

Radio Frequency Chokes (RFC)

Although, I'll also wind, or perhaps ply carefully dimensioned copper board traces to make RF chokes, I built up a stock of lower Q chip inductors in size 0805 and 0603. These include 8.2 nH, 33 nH, 68 nH, 220 nH, 330 nH, 470 nH and a few other values. For RFCs we need to consider L, tolerance, SRF, Q , maximum current and temperature stability — it's probably better to only buy clearly specified parts so you can study their datasheet to glean this information. I've swept nearly all my RFCs and recorded their SRF.

Choosing RFCs — SRF (series resonant frequency)

An RF choke exerts its greatest impedance at its SRF and in the case of a first-order RFC such as a single inductor, we should aim the SRF at or near the frequency we wish to attenuate the most. Consider a choke in the emitter lead of a BJT Colpitts oscillator with a frequency range from 375 to 400 MHz: The choke's ideal SRF should lie somewhere above 400 MHz. As possible, the SRF of a first-order RFC should exceed the operating frequency of the device it's choking.

In higher-order filter applications such as a C-L or C-L-C (pi filter) applications, engineers tend to target the choke's SRF to >= 10 times the operating frequency. Challenges such as limited inductor choices, needing to match stages, or coping with stray reactances will arise, however, as amateur UHFers, we just try our best to fit an RFC.

UHF Capacitors

I swept numerous size 0805 + some 0603 and 1206 capacitors in a microstrip line jig to determine their SRF. Math formulas enable us to calculate and plot attentuation, Q and ESR at self-resonance if we wish. Click for 1 example. Although we rarely apply capacitors at their SRF in practical circuits, considering Q, ESR and attenuation gets us thinking that capacitor SRF is a dynamic value that may vary widely with part size, capacitance and datasheet specified Q.

For instance, if you take a size 1206 capacitor, measure its SRF and then replace it with a same-value size 0805 capacitor, the SRF of these two caps might vary by as much as 200 MHz or more. Size does matter! Even the same value capactor by a different manufacturer may be a SRF game changer. Bob, K3NHI tells it correctly: to measure *is* to know.

Datasheet specified capacitor Q is usually measured at a low frequency like 1 MHz. Since Q decreases as frequency increases, a capacitor that specs a Q of 2500 at 1 MHz may only sport a Q of 87 at its SRF way up at 512 MHz. During sweeps, I measured a deeper attenuation notch and often a higher SRF for "very high Q" caps when compared to the common, low-medium Q, popcorn caps sold as mystery parts by online stores or auctions. I'm now hypervigilent about finding very high Q SMD capacitors for sale at a low price.

I wrote the SRF of all my SMD caps in my notebook and use this data to inform my capacitor choice for RF bypass and tuning. Click for a photo of 3 UHF parts drawers. I've pretty much gone to size 0805 caps to try and garner a higher SRF at VHF and UHF. Still, too, we might have to solder down 0603 parts up above 1 GHz. Yikes!

Did you notice that when you drop a 0805 part on the floor, you never ever find it again? Even a little cough can expunge a size 0603 part from your QRP workbench.

50 Ω Detector

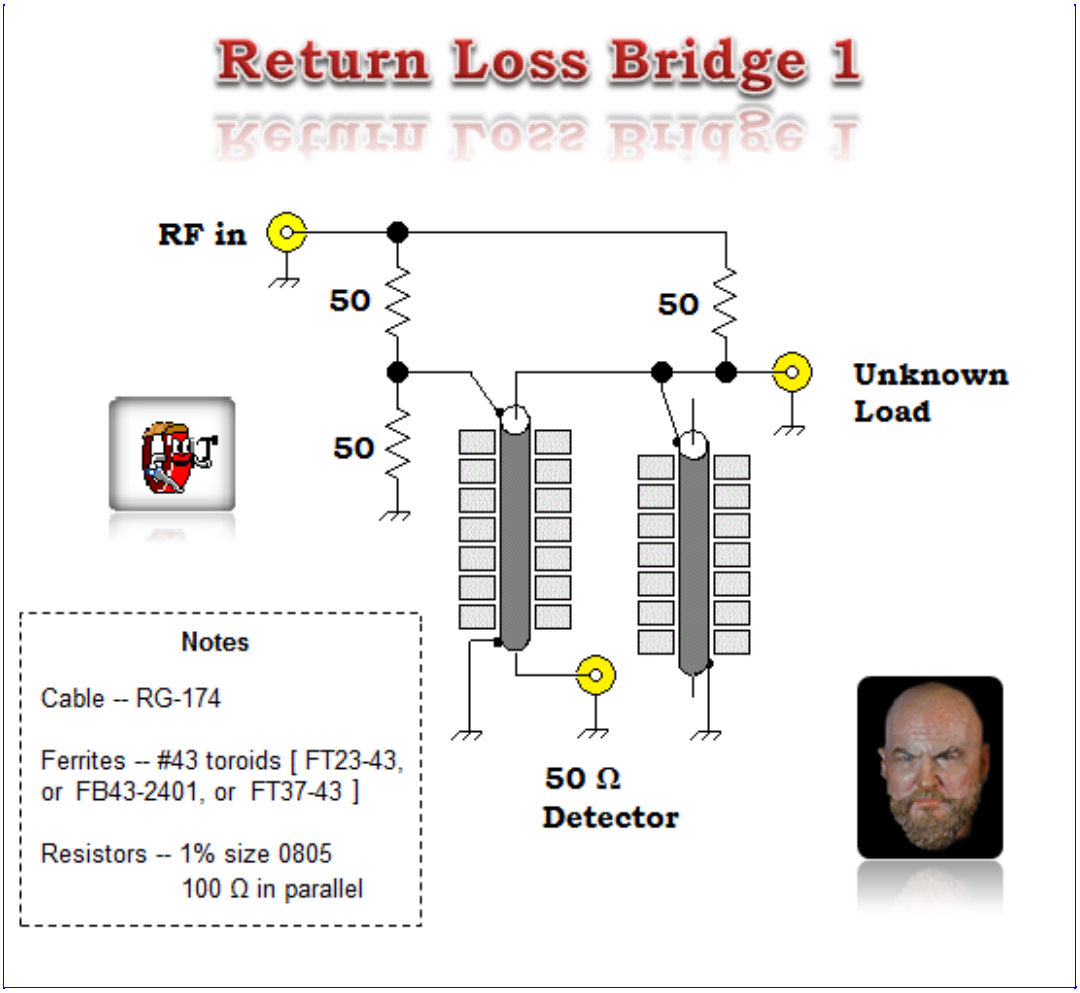
From 0.5 - 200 MHz, my 50 Ω terminated input DSO forms the heart of my test bench. An analog, or digital storage oscillocope with a bandwidth >= 1 GHz, plus a suitable wideband probe costs dearly and for many, won't be an option. UHF bandwidth, thermistor-based 50 Ω power meters are also expensive and lack signal viewing.

Thus, a spectrum analyzer seems the best choice for the amateur UHF enthusiast. A spectrum analyzer with a built-in tracking generator adds serious value — giving us a MF-UHF sweep system with profound utility. One popular, modern TG + SA choice is the Rigol DSA-815. I'm fortunate to have access to gear that measures up to 3 GHz.

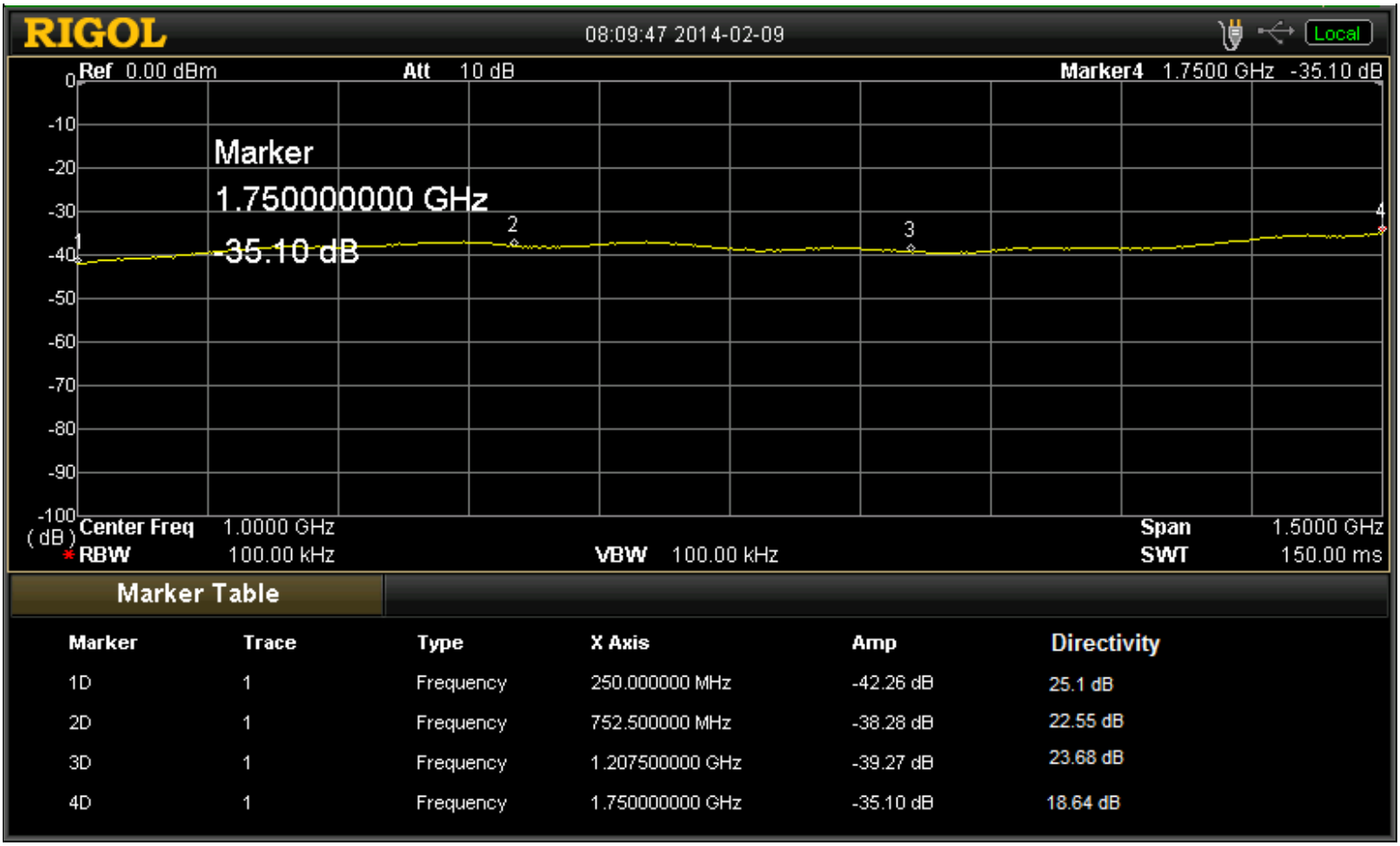
3. Return Loss Bridge Experiments (RLB)

I'd be lost without a return loss bridge at UHF. Directivity marks the touchstone of a return loss bridge — and up at UHF grabbing a minimum 20 dB directivity may pose quite a challenge. Experiments to find a bridge with good directivity from 250 to 1750 MHz follow:

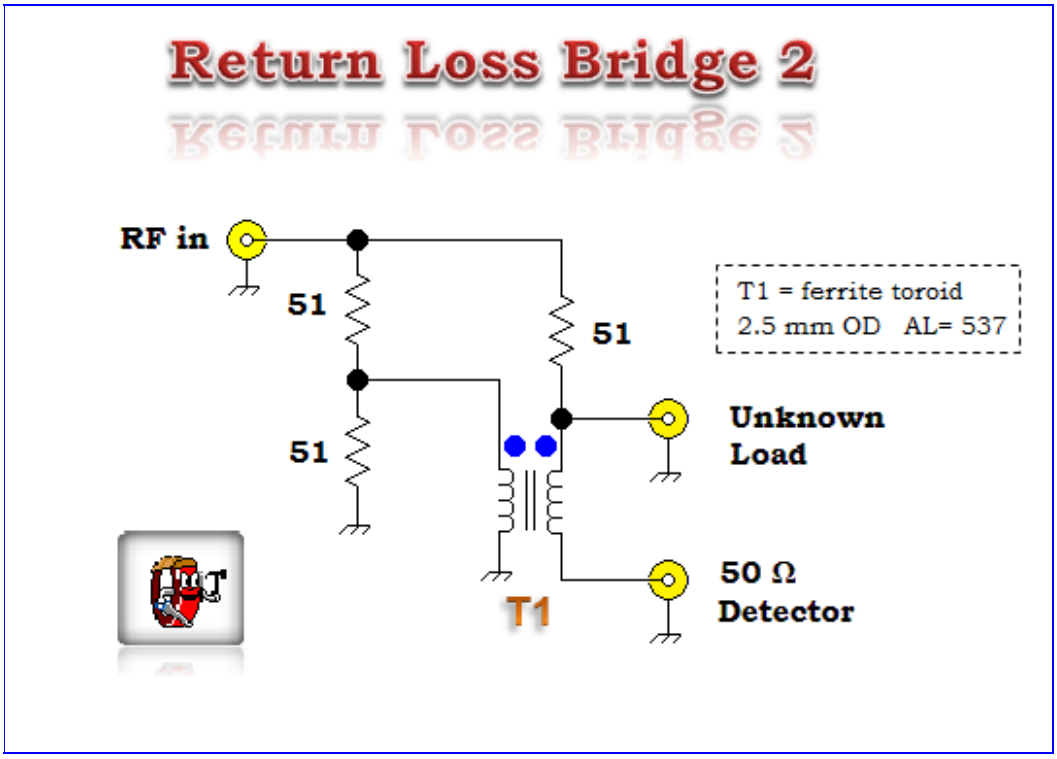
Above — A macro photo of RLB#1: 1 of the 3 "string of pearls" type return loss bridge built this Winter. I tried different numbers, sizes and mixes of ferrites. This particular bridge had 3 and then 2 FT37-61 toroids as shown, although I did not shorten the RG-174 coax when reducing to 2 toroids. The 61 mix gave poor directivity and was abandoned.



Above — Return loss bridge #1: the "string of pearls" variant. Several web sites feature great "string of pearls" RLB experiments for you to review and many place enough ferrite to extend from HF to UHF. I swept each of my builds in a Rigol tracking generator + spectrum analyzer with 4 frequency markers set. I made an open-circuit sweep [nothing on the unknown impedance port] followed by a sweep with a Mini-Circuits Labs DC - 18 GHz, 50 Ω load threaded on the unknown Z port. The difference between the open-circuit and 50 Ω measurement = the directivity at that frequency.

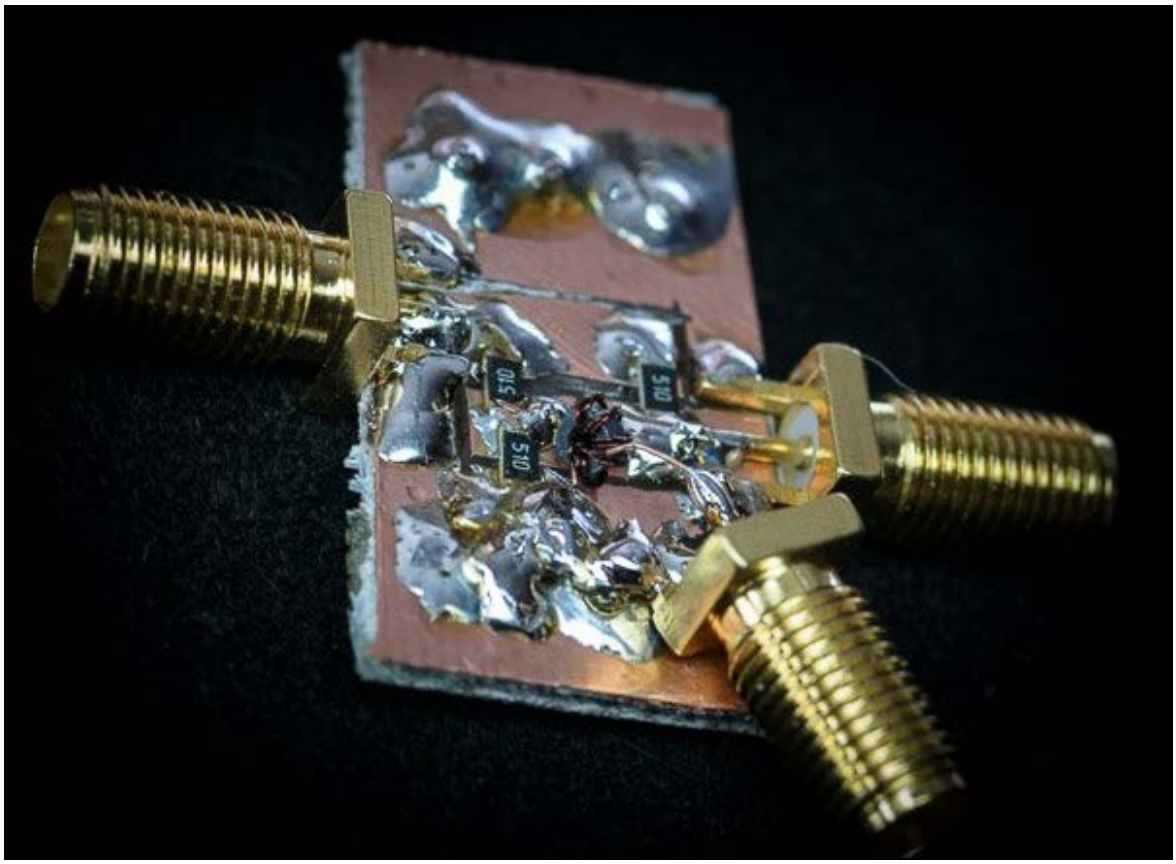


Above — The sweep of my best "string of pearls" bridge from 250 MHz to 1.75 GHz showing the directivity at 4 frequency points. Click for the open circuit sweep. I felt encouraged by this sweep with five FB43-2401 beads over each piece of coax cut just long enough to fit the beads. I built on 2-sided FR4 copper clad board with at least 6 via wires connecting the 2 copper surfaces.

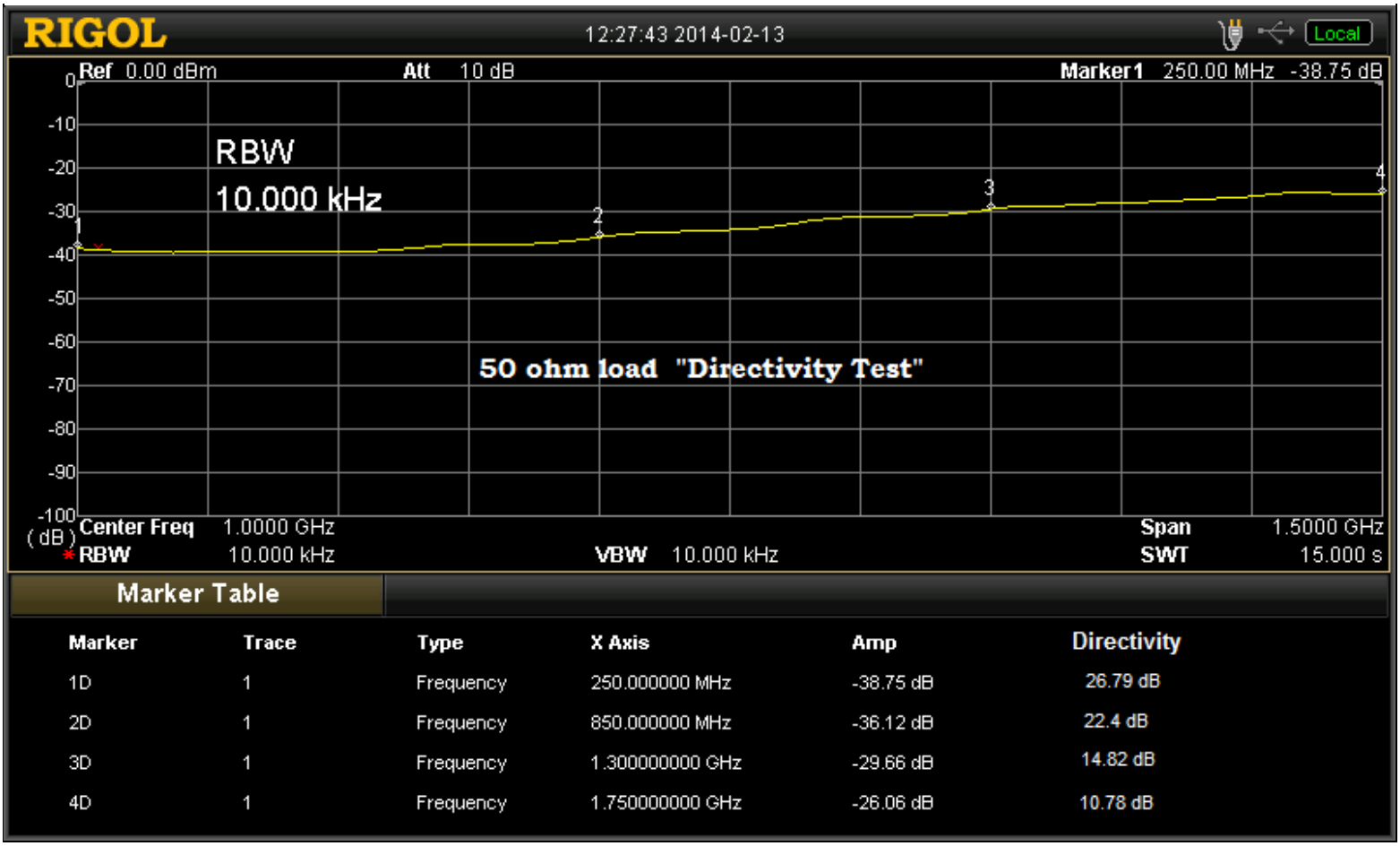


Above — Schematic of RLB #2: the bridge topology featured in EMRFD. T1 = 5 bifilar windings of 32 gauge wire wound on a *tiny* ferrite toroid boasting a strong AL. I pulled this ferrite from a blown up mixer removed from an old 1.5 GHz spectrum analyzer. I started with 5 toroids, but during winding, got spastic and dropped 2 of them on the floor and concluded that the SMD part gnome who lurks beneath my bench swallowed

them up, since I couldn't find them after an intensive search.



Above — Magnified return loss bridge #2 breadboard. Serious UHF enthusiast Paul Wade, W1GHz found that the smaller your breadboard, the better the UHF directivity in his bridge experiments: My findings support his. In RLB #2, I went with plain, size 1206, 5% resistors. I just ordered some size 0603 49.9 Ω 1% resistors for future experiments in my quest to make my ultimate UHF RLB.



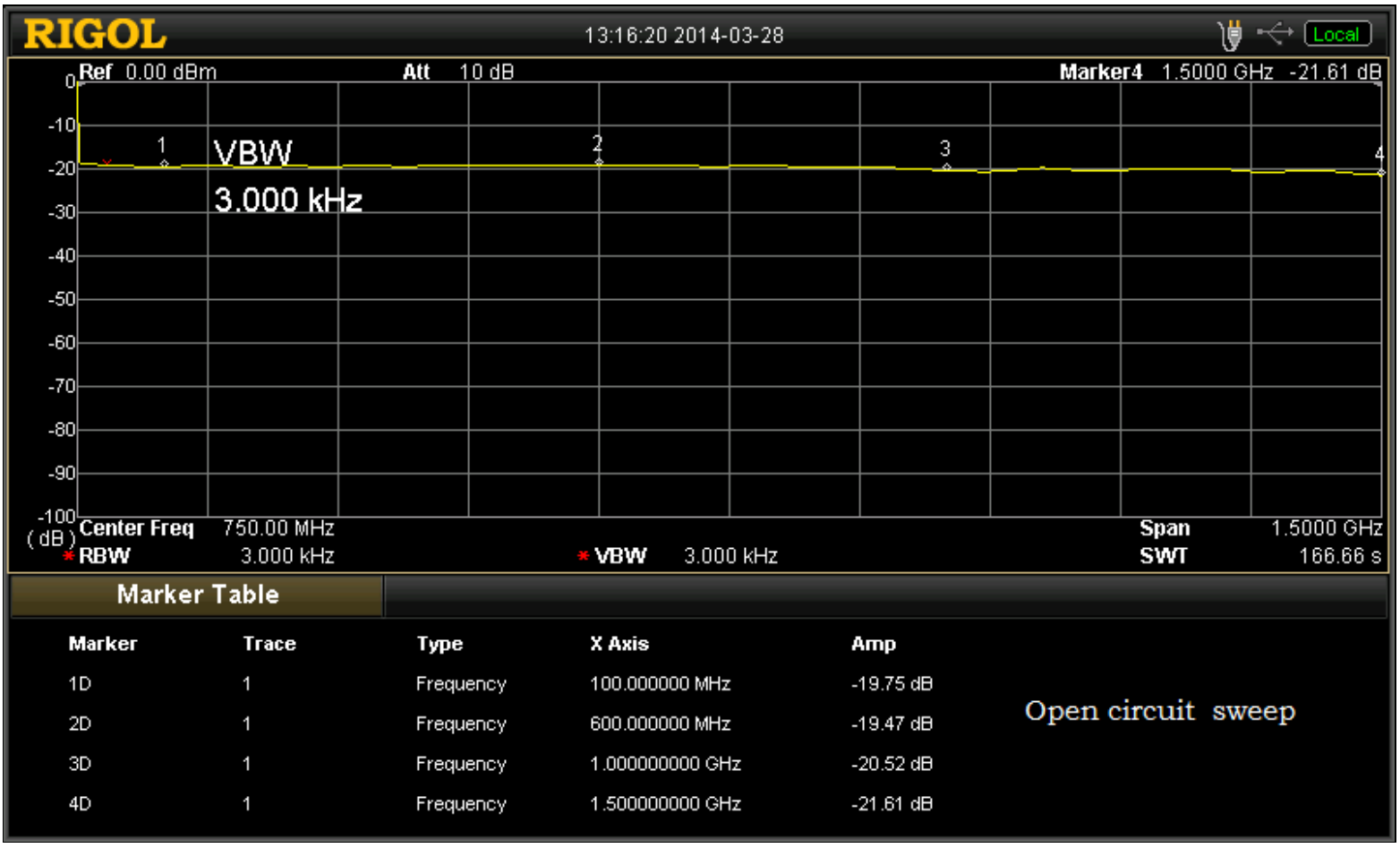
Above — The sweep of RLB #2 from 250 MHz to 1.75 GHz showing the directivity at 4 frequency points. Above ~900 MHz, directivity falls right off, so this bridge won't cut it for a serious UHF bench. I tried reducing T1 to 3 bifilar windings but this worsened the directivity. Click for a different sweep of RLB # 2 from 50 MHz to 1.5 GHz. Directivity = 27.5 dB at 50 MHz, so this bridge would work great for VHF. I imagine that SMD 49.9 Ω 1% resistors might boost directivity a little.

Of course, we're always impressed by the work of Kostas , SV3ORA.

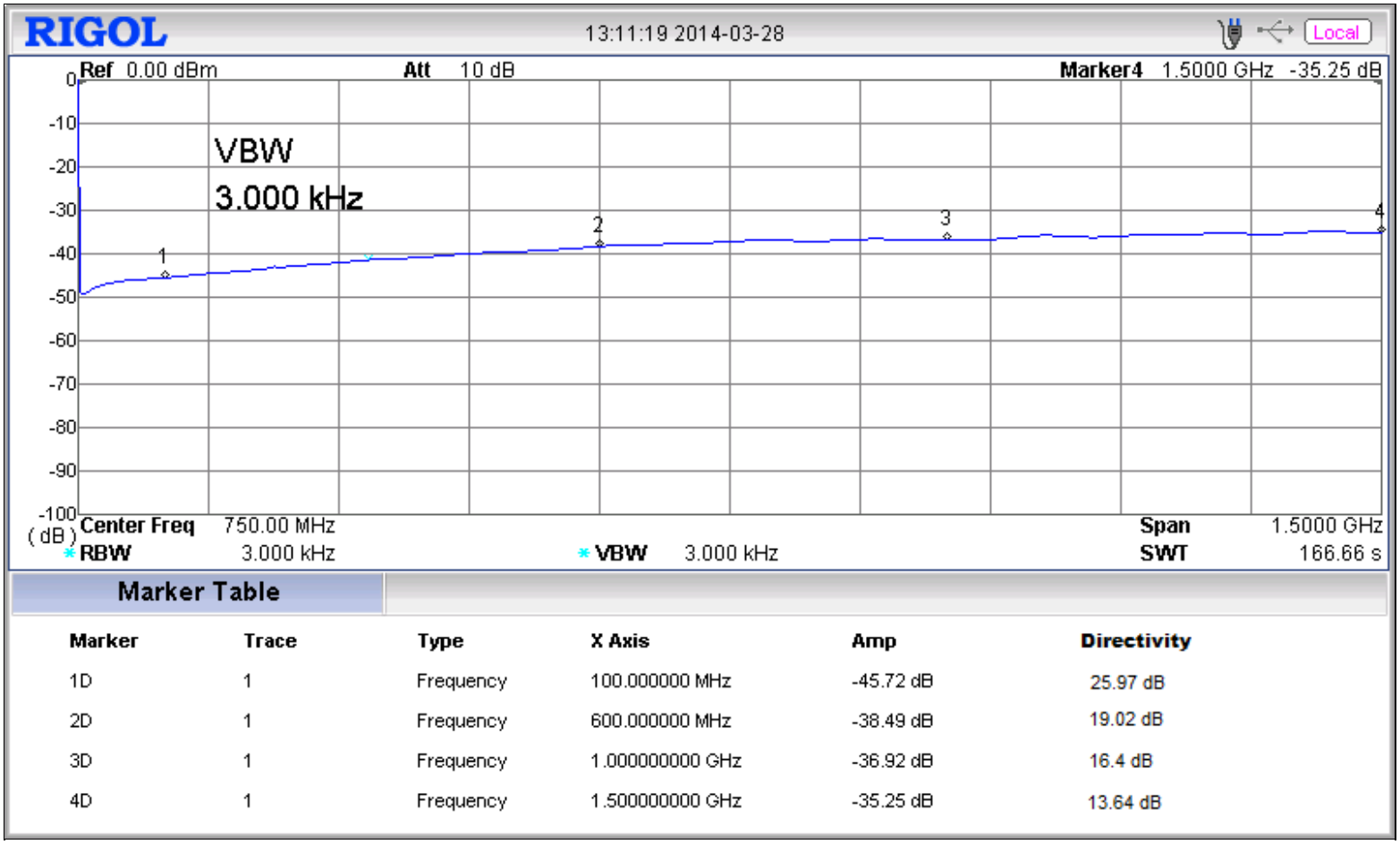
Return Loss Bridge 3

Some experiments with the 3-bead RLB from Sam Wetterlin follow:

Above — My build of the Wetterlin 3-bead return loss bridge. This PC board was the only 1 left over from a 2006 joint project between the 4 Hams whose calls are etched on the PCB. Wes sent me this board and some of the parts and I'm grateful. My 3 dB pads = 294 Ω 1% resistors plus an 18 Ω 5% in size 0805. My 49.9 Ω 1% resistors = size 0603. The transformer = a Minicircuits Labs TC1-1-1-13M+. Click for the ferrite description. I used 24 gauge copper via wires to connect grounded nodes to the ground plane on the back of the PC board. I chose to permanently solder a 49.9 Ω resistor on the reference port.



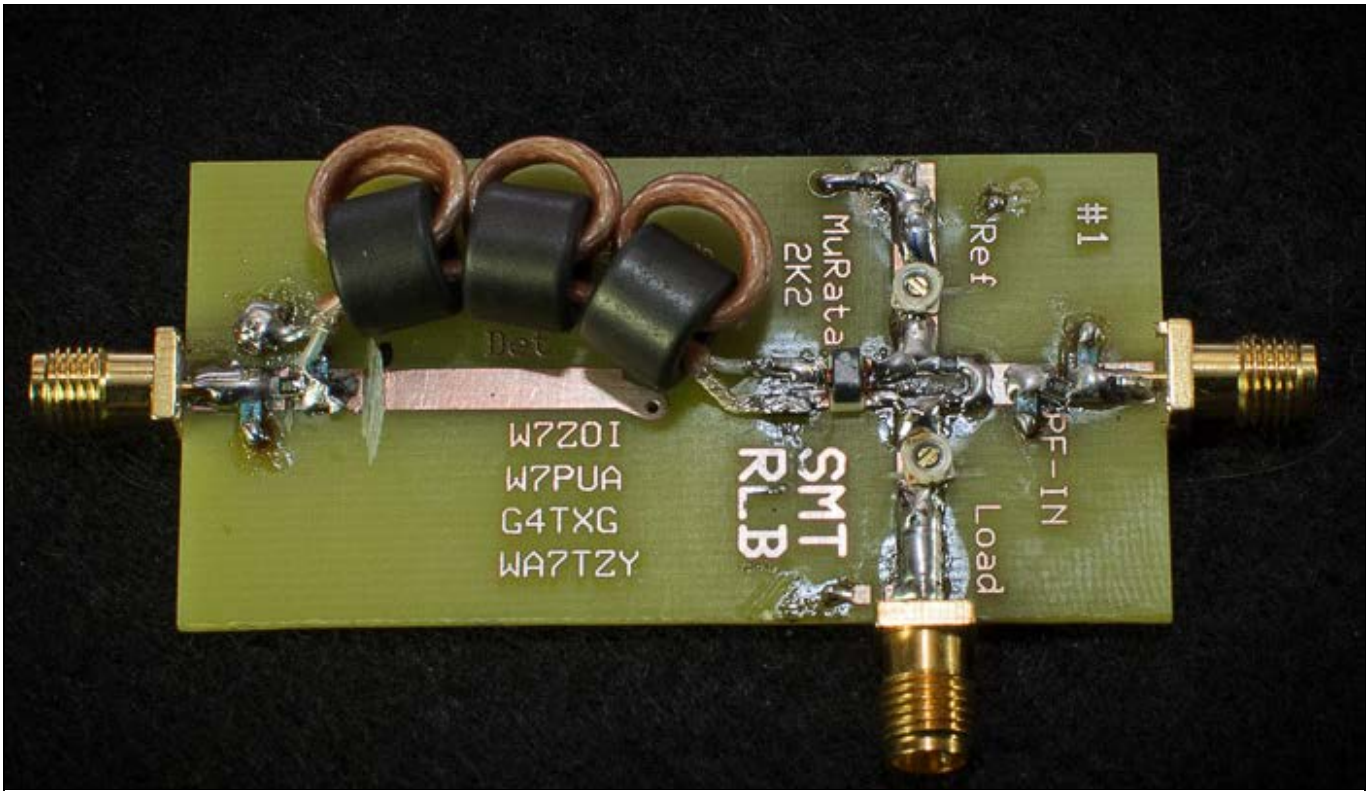
Above — Open circuit sweep with 4 markers set. Span = 1.5 GHz.



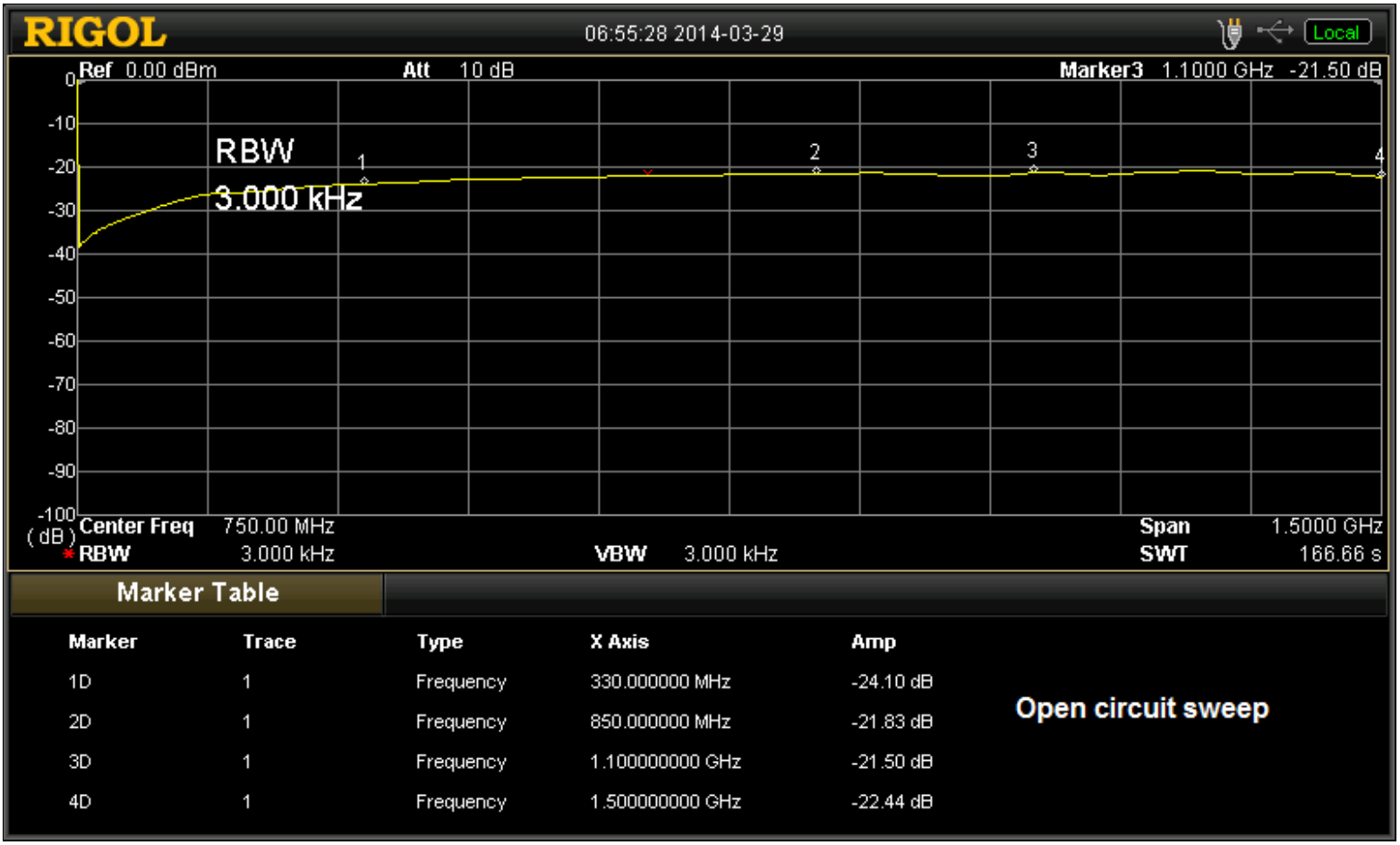
Marker Table

Marker	Trace	Type	X Axis	Amp	Directivity
1D	1	Frequency	100.000000 MHz	-45.72 dB	25.97 dB
2D	1	Frequency	600.000000 MHz	-38.49 dB	19.02 dB
3D	1	Frequency	1.000000000 GHz	-36.92 dB	16.4 dB
4D	1	Frequency	1.500000000 GHz	-35.25 dB	13.64 dB

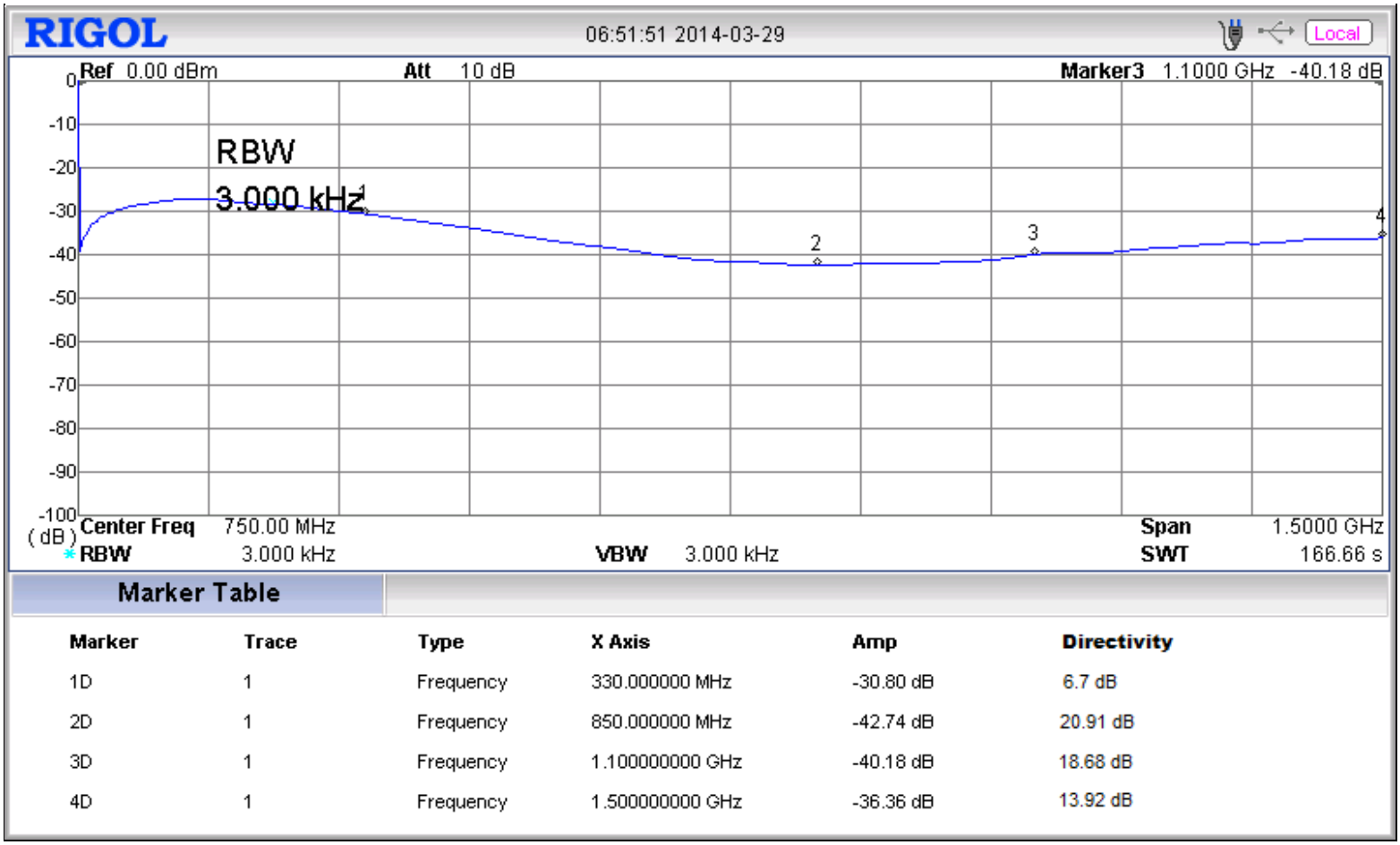
Above — 50 ohm load termination sweep with calculated directivity. Click for the 50 Ω sweep where the span = 200 MHz. I was disappointed with the UHF directivity in *my* build.



Above — The 3-bead RLB with added series trimmer capacitors on the load and reference ports. SMT trimmer caps = Voltronics JR300 5.5-30 pF with a Q of 1500. I tried both shunt and series caps to try and boost UHF directivity.



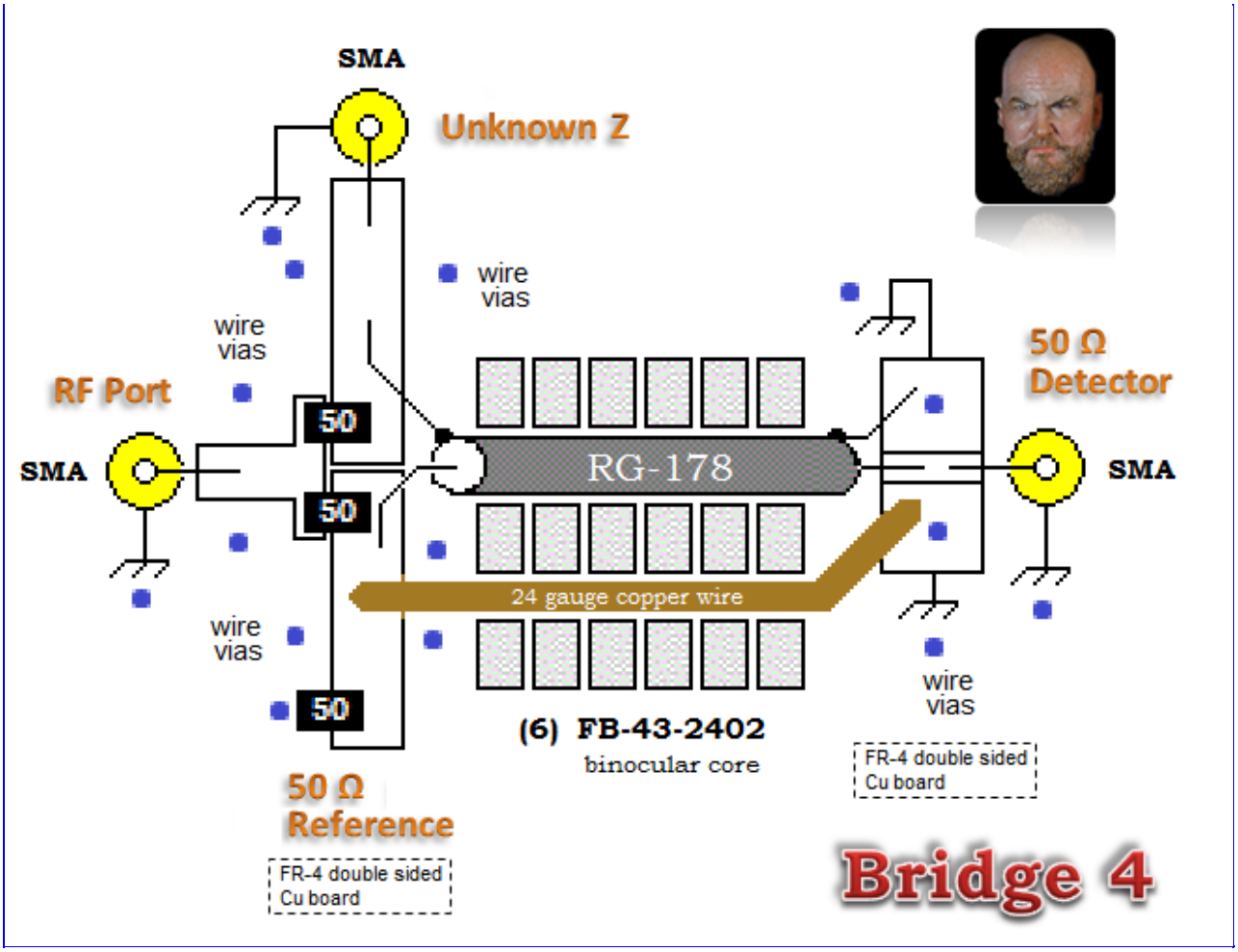
Above — Open circuit sweep of the 3-bead bridge with series capacitors on the load and reference ports.



Above — 50 Ω load sweep. By carefully tweaking the 2 trimmer caps, I set the deepest NULL at ~ 850 MHz; although it's not as deep as I want. Low end directivity was compromised.

Return Loss Bridge 4

Since my bench experiments with the "string of pearls" (SOP) return loss bridge showed promise, I kept going and eventually built a version closer to my goal: A bridge for 300 MHz to 2 GHz with decent directivity.

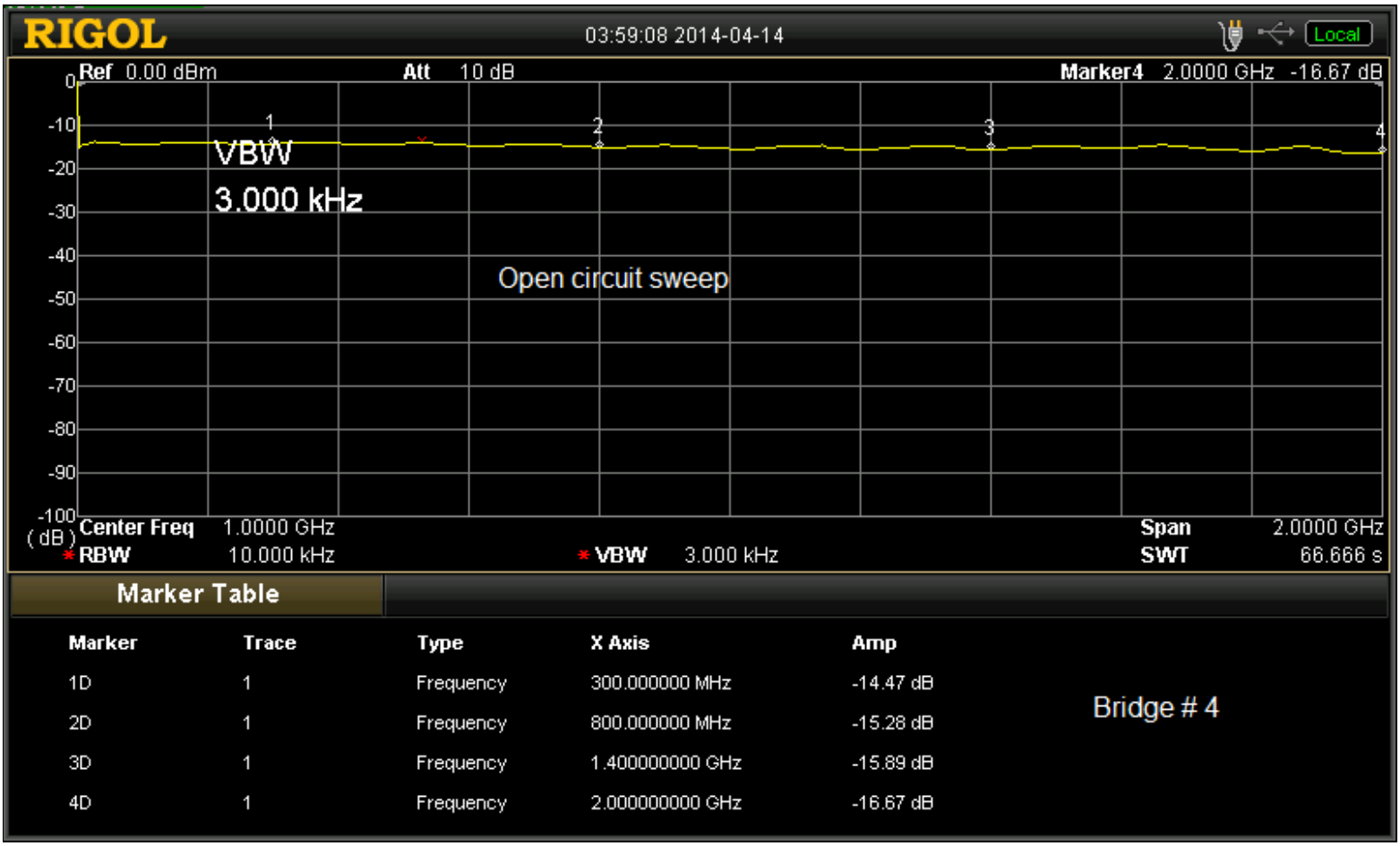


Above — My SOP build diagram of Bridge #4. I cut traces in 2-sided FR-4 Cu board to make the needed microstripline paths.The two 49.9 Ω 1% bridge resistors = size 0603, however, my reference port 49.9 Ω, 1% resistor was size 1206. The braid of the coax goes to the unknown impedance port and the center core to the reference port. Six FB-43-2402 binocular cores just passed the diameter of the RG-178 coax and binocular ferrites boosted balance compared to seperate beads in my experiments.

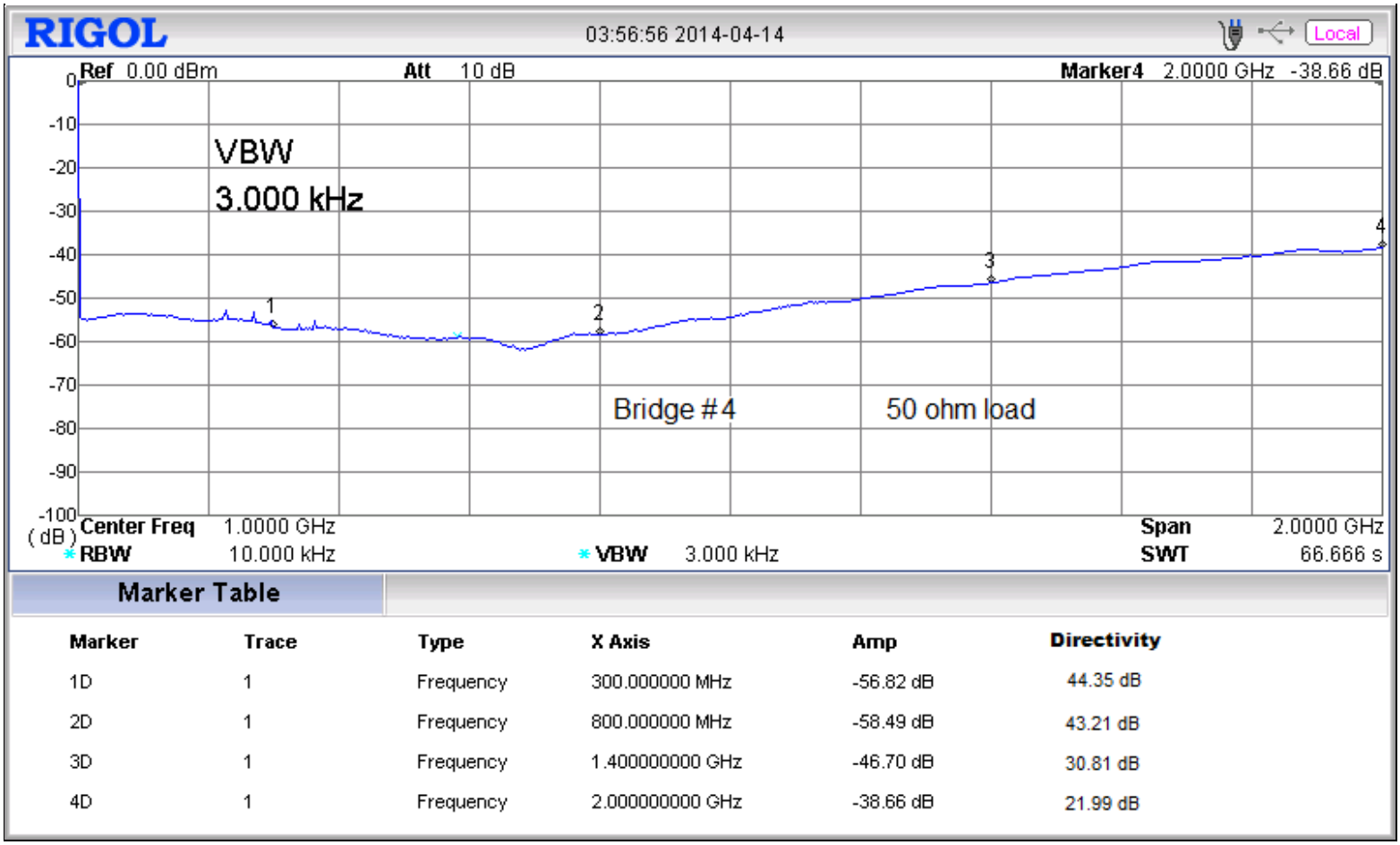
Rather than run coax through the remaining hole in the ferrite string, I just used a 24 gauge piece of Cu wire for balun symmetry. I left the green insulation on the wire to help keep it in place during soldering — I used the same wire spool to make the via wires.

Above — Breadboard photograph. Click for a larger photo. I soldered some thin brass sheet metal to electrically join the ground planes of the 2 carved boards [represents the normal RF-tight metal chassis we apply].The brass sheet boosted directivity 8 dB by improving balun function.

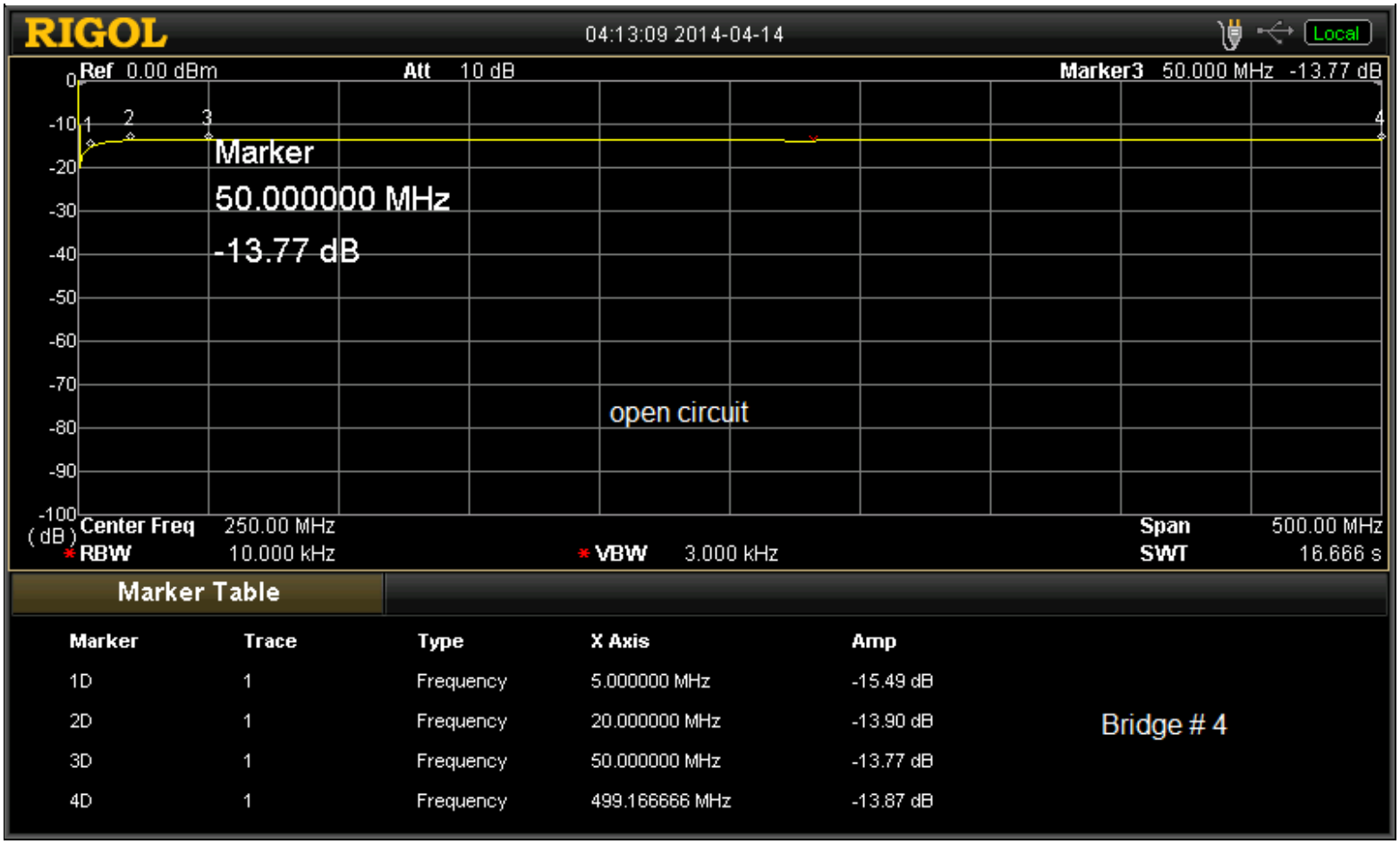
Comments by a prominent figure in our hobby ....."and the thing is 'ugly to the core'. Way to go."



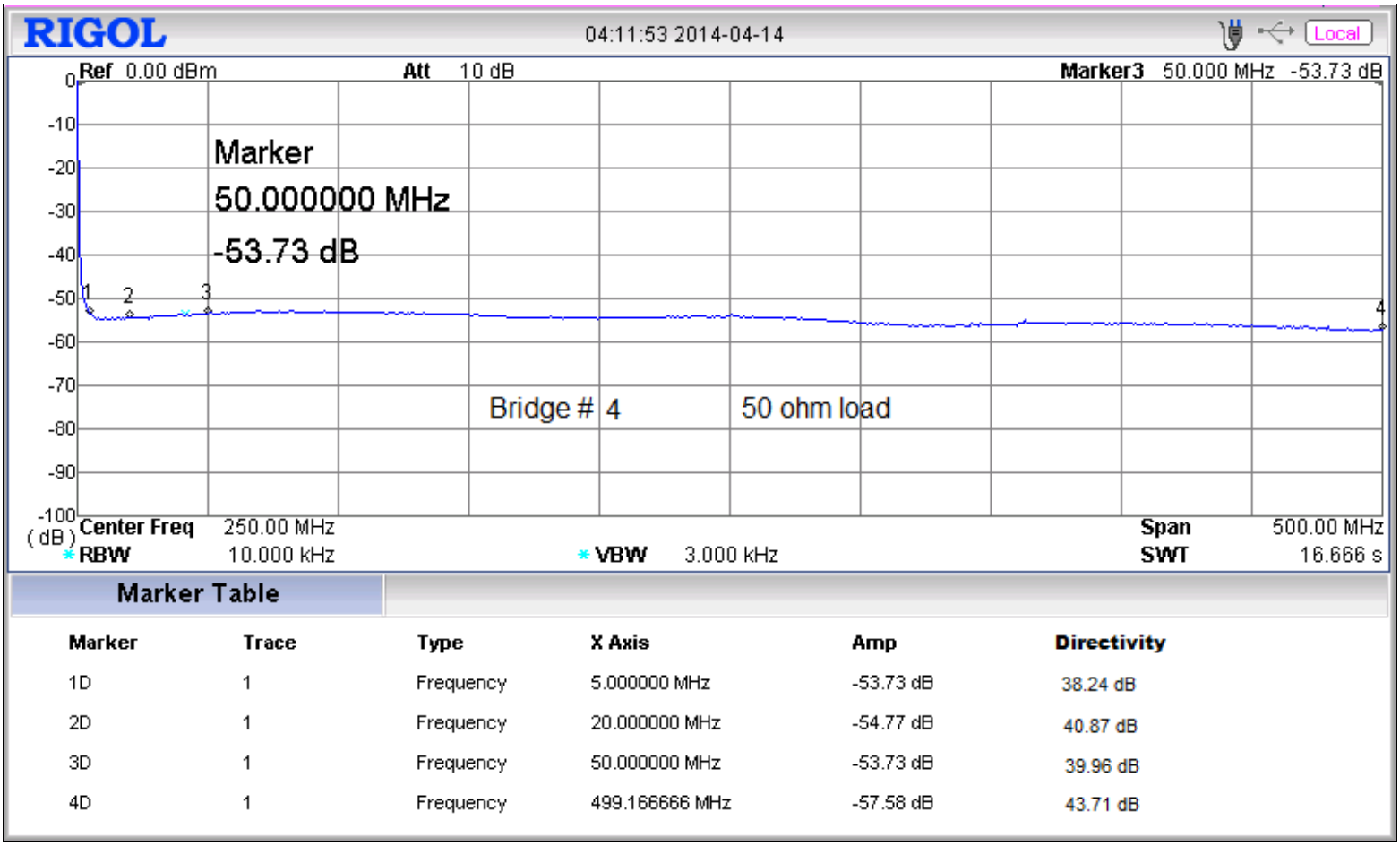
Above — Open circuit sweep with a 2 GHz span and center frequency of 1 GHz.



Above — Sweep with a 50 Ω load threaded on the unknown impedance port and 4 markers showing directivity at those 4 frequencies. Yay — this is much better! I wanted at least 30 dB directivity at 1296 MHz and got it.



Above — Open circuit sweep with a 500 MHz span and center frequency of 250 MHz. I set 2 of the markers to evaluate the directivity at HF.



Above — 500 MHz sweep of Bridge #4 with a 50 load on the unknown impedance port. The directivity at HF surprised me. Although better suited for VHF-UHF, this bridge will work okay from 5 MHz to ~1.5 GHz.

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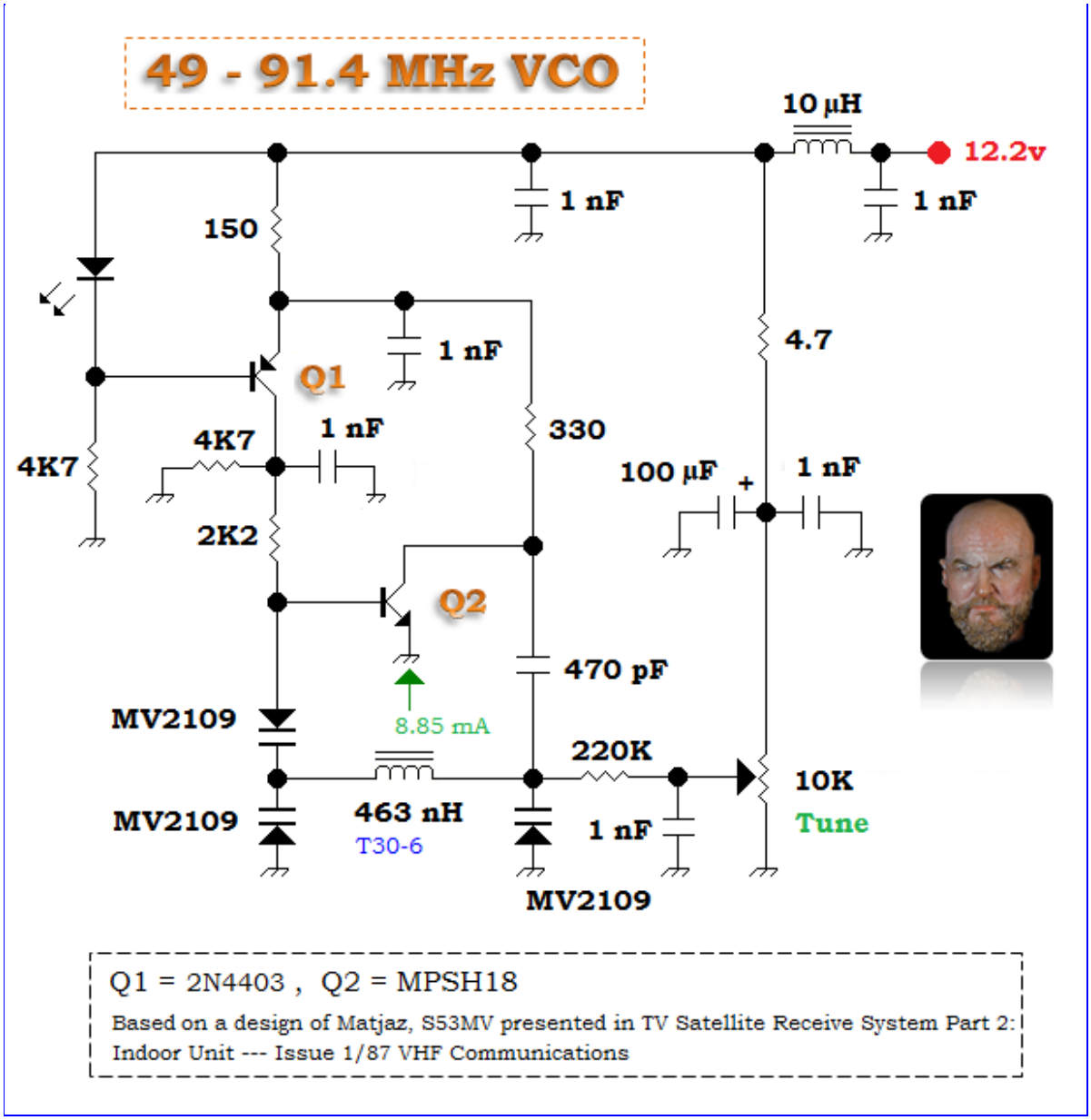
4. VCO Experiments

Ask experimenters about homebrew RF oscillators and you might hear about a pleasing paradox — vexing and satisfying in the same breath! At UHF we usually make a VCO and keep it on frequency by phase/frequency locking to a xtal oscillator reference. I sought some wideband, free running VCOs in hopes I'd inch up the learning curve.

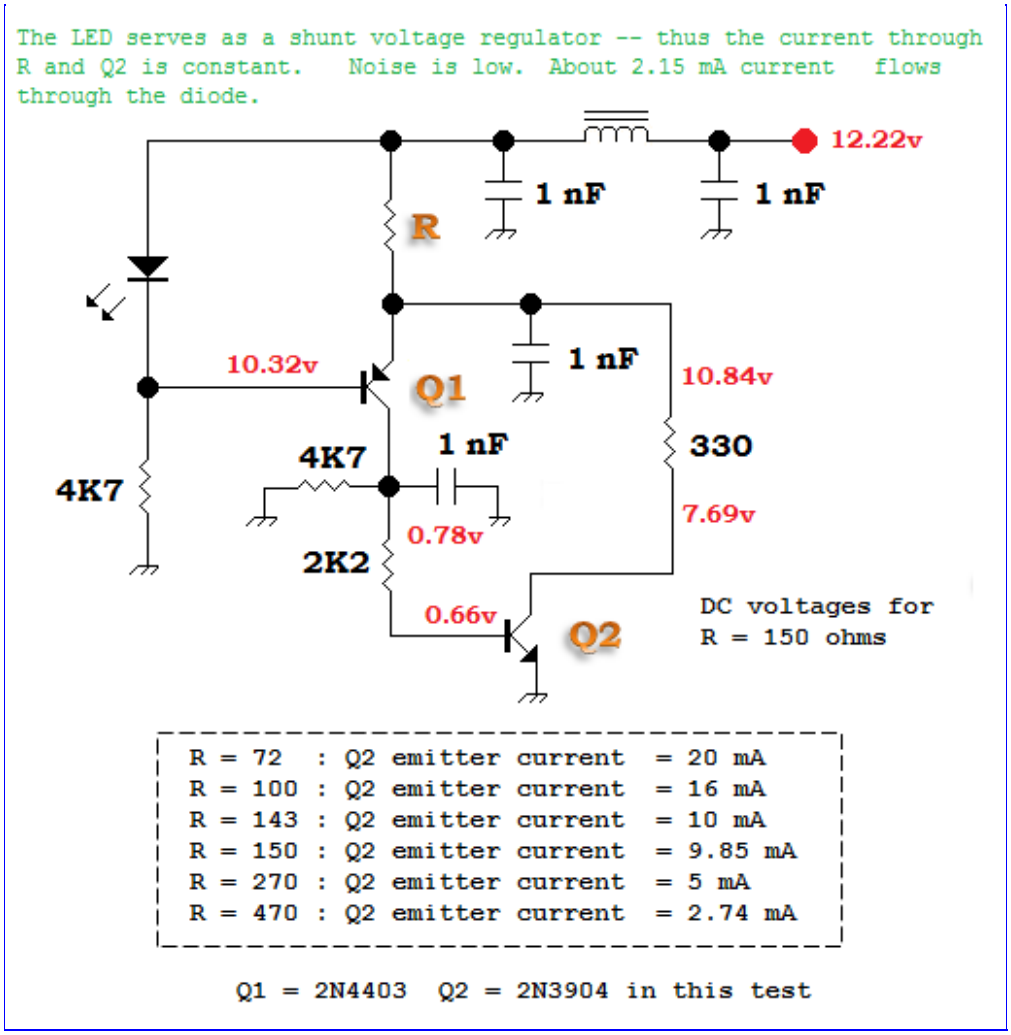


Above — My first attempt at a VCO circuit described by Matjaž, S53MV in a 2-part article published in VHF Communications magazine. The original circuit contained a tunable RF filter, mixer, buffer amplifer + more and I recommend you purchase the 2 articles for your own analysis. Shown with permission from Andy Barter of VHF Communications magazine, I ordered all issues from the 1980s which came on this DVD.

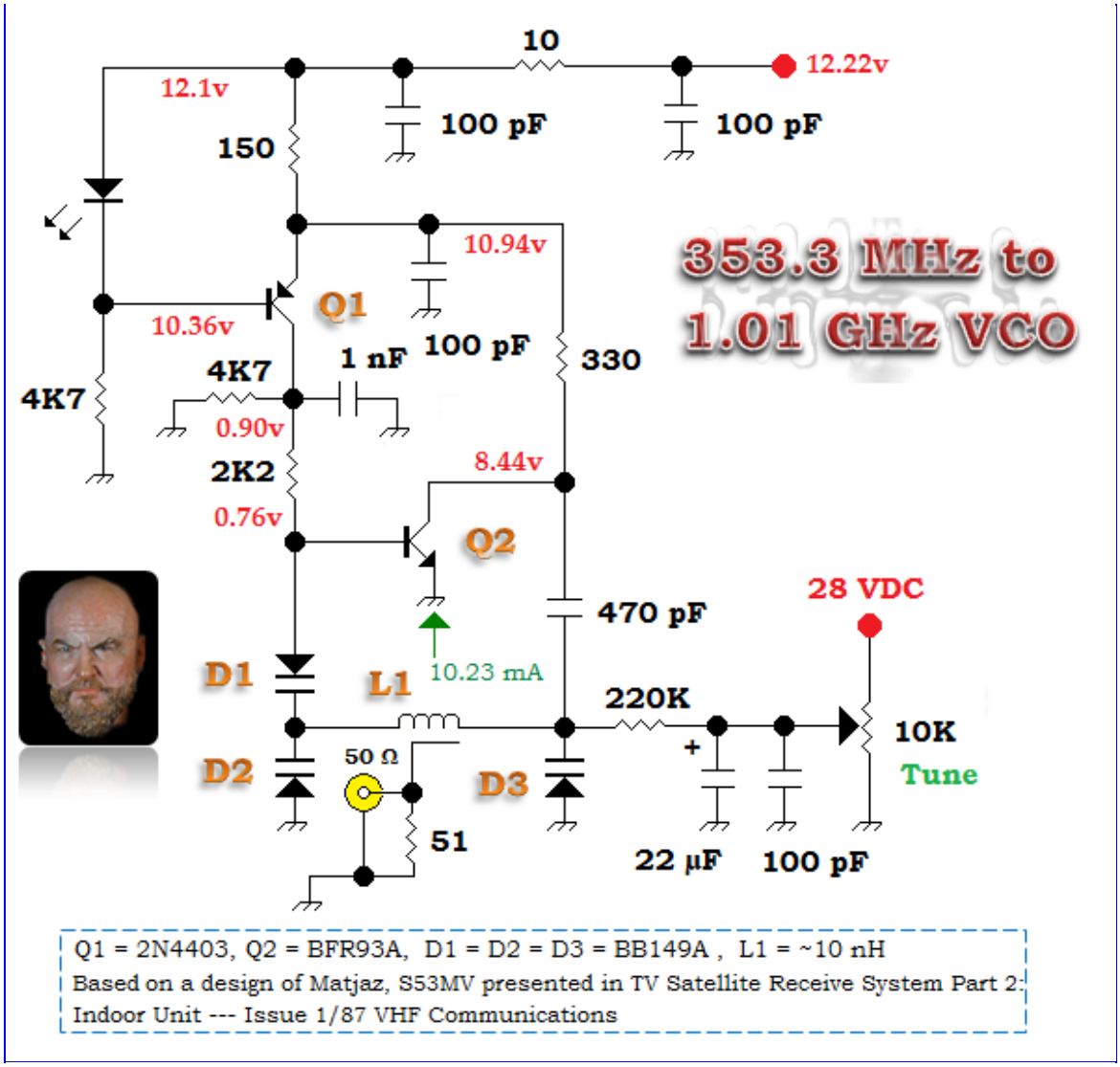
With 28v varactors, this VCO will tune 1 octave minimum.

To bench investigate this stunning common emitter Colpitts variant, I first built it at VHF using through-hole parts with a standard 12 volt tuning voltage that tuned from 49 to 91.4 MHz. The output came from 1 link around the 463 nH inductor terminated in a 1K resistor at the hot end + ground at the cold end to allow 10X probe measurement with my DSO.

Matjaž, S53MV describes his circuit function here. A variable C on the base in combination with the input capacitance of Q2 operates as a tuned impedance matching network and preserves a steep phase slope (which means you have higher Q) — a requirement for low phase noise. In essence, the inductor with a varactor at each end forms an adjustable pi-network to provide correct phase shift tracking.



Above — I built the low-noise, temperature-stable, PNP current source using though-hole parts. R provides a convenient way to vary collector current, although changing the PNP base bias also works.



Above — My UHF VCO schematic. Click for the breadboard photo. A wire positioned near L1 provided the output into my spectrum analyzer. Click for a sweep with the reverse DC voltage set at 0v. Click for the a sweep with the tuning at 28v to establish the lowest possible varactor capacitance. Markers indicate the carrier frequency. As shown, a common emitter VCO exhibits rich harmonic energy and viewing the second harmonic at 2 GHz felt really cool in the 28v sweep. Not to worry — the tracking generator in the Rigol TG + SA units also exhibit strong harmonics and work fine for sweeping.

Click for a pdf file showing how I laid out my board. I cut traces on the top of 2-sided FR4 board and connected any grounded points on top to the bottom ground plane with a 24 gauge copper via wire. A 28v DC-DC Converter to supply the varactors lies on VHF-FM.

I varied L and even tried a 3 mm by 8 mm copper sheet metal inductor but this failed to oscillate. In the end, a single wire with ~10 nH inductance worked to give me the tuning span shown. L1's distance above the ground plane was determined by tuning the VCO for the smoothest transistion across its range. This circuit begs further experiments including Q2 current, L1 and the 470 pF fixed feedback capacitor.

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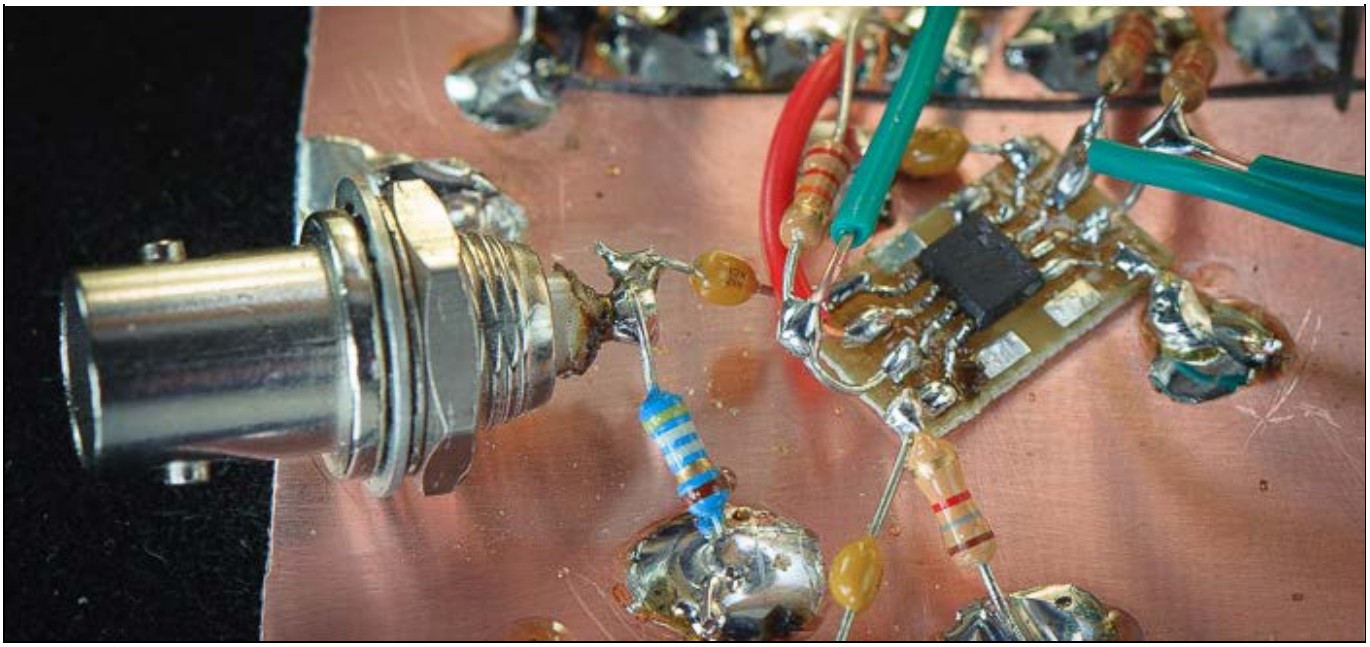
5. Frequency Counter Kit

A sensitive UHF bandwidth frequency counter makes an important tool for the UHF homebrew experimenter. With a search engine, you'll find commercial and kit counters for sale along with DIY projects for builders handy with microcontrollers.

If you already own a decent counter, then a crude prescaler tool employing an IC like the MC12079 might work okay if you can tolerate or manage modulus division by 64, 128, or 256. The MC12079 requires an RF signal between -17 dBm to +3 dBm to count properly, so many builders place amplifier(s) along with anti-parallel clamping diodes before the digital IC. The Motorola datasheet shows a basic circuit to build off.



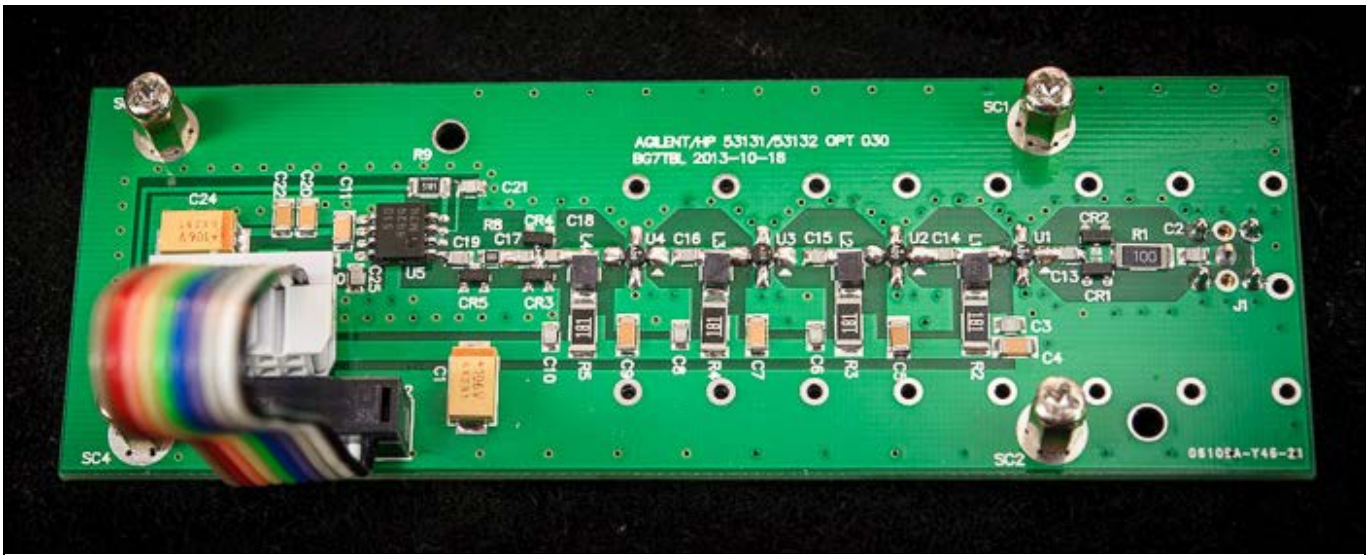
Above — My first UHF counter experiment involved the Motorola MC12079 and 12080 prescalers. Eventually I made a proper double-sided copper board device with anti-parallel diodes, a MMIC, the MC12079 and some 74AC series logic — it worked okay for casual use. The 2 big drawbacks were MC12079 self-oscillation with 0 applied input signal and having to remember to multiply measures by my set prescaler division integer. I sought a long-term solution.

A Solution Appears

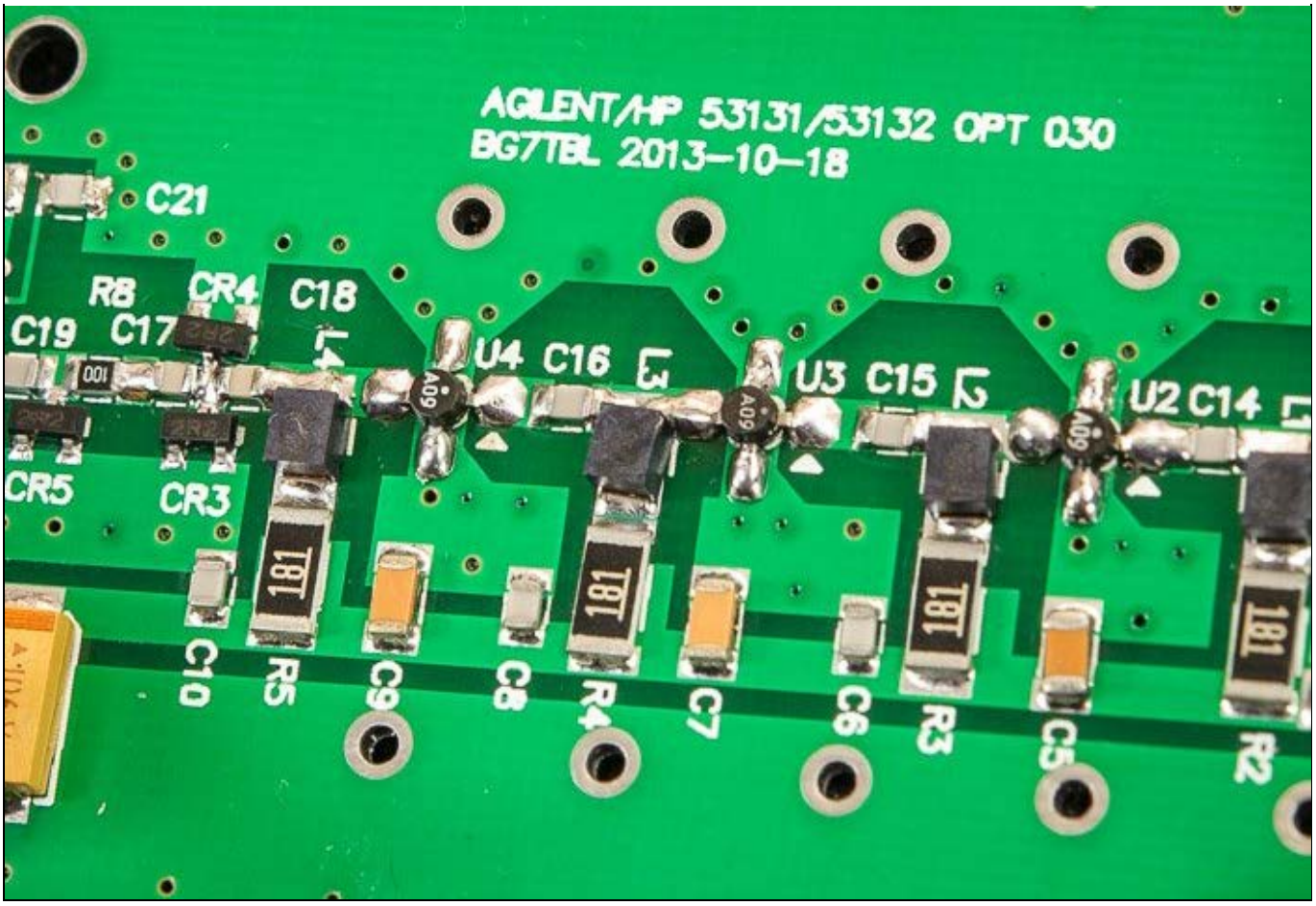
In 2012 I bought a second-hand, calibrated HP 53131A with the optional A6 High Stability Timebase from an EE for ½ the going price at the time. 225 MHz bandwidth poses its only limitation; it even came with the original spiral bound manual.

This summer, Shahriar from TheSignalPath.com reviewed a Chinese kit to take the HP 53131A to 3 GHz with a clone of the original HP OPT30 Channel 2 kit. Based on his quantitative review and warm thoughts of using my sturdy HP counter at UHF, I got 1. Mine sold on eBay for $87 Canadian including shipping. Click for TheSignalPath.com video. Click for a larger photo of the kit I bought.

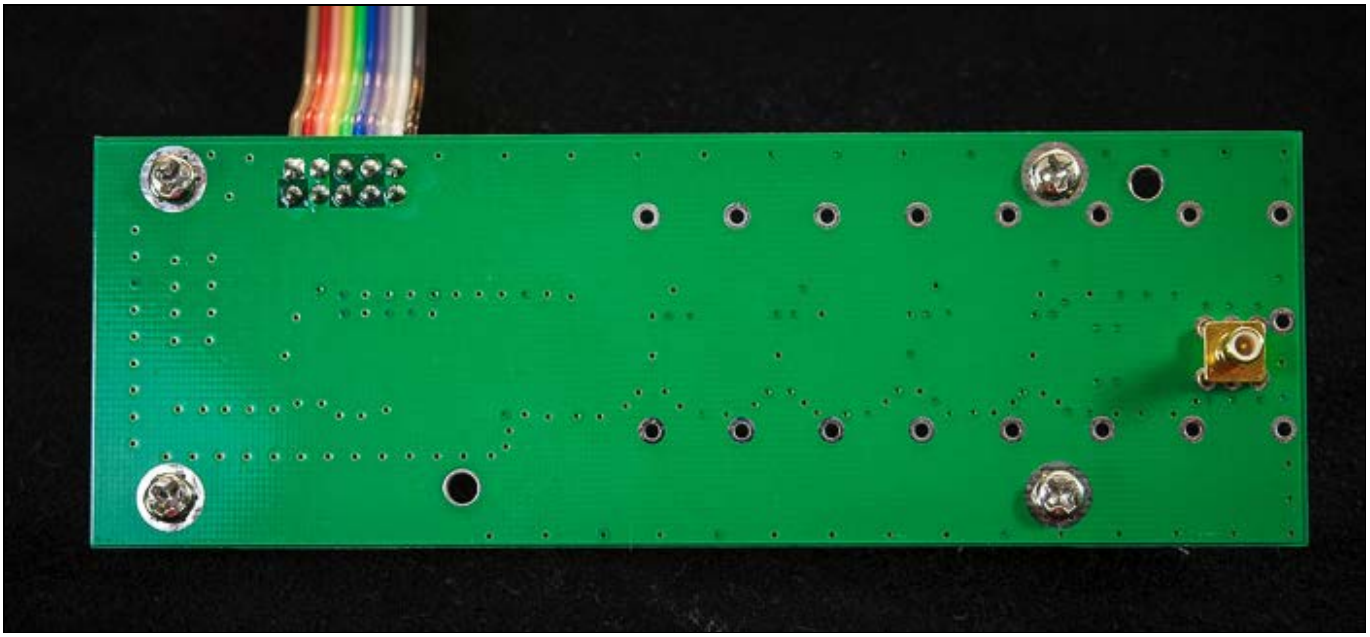
Above — My HP counter with the kit installed as Channel 2.



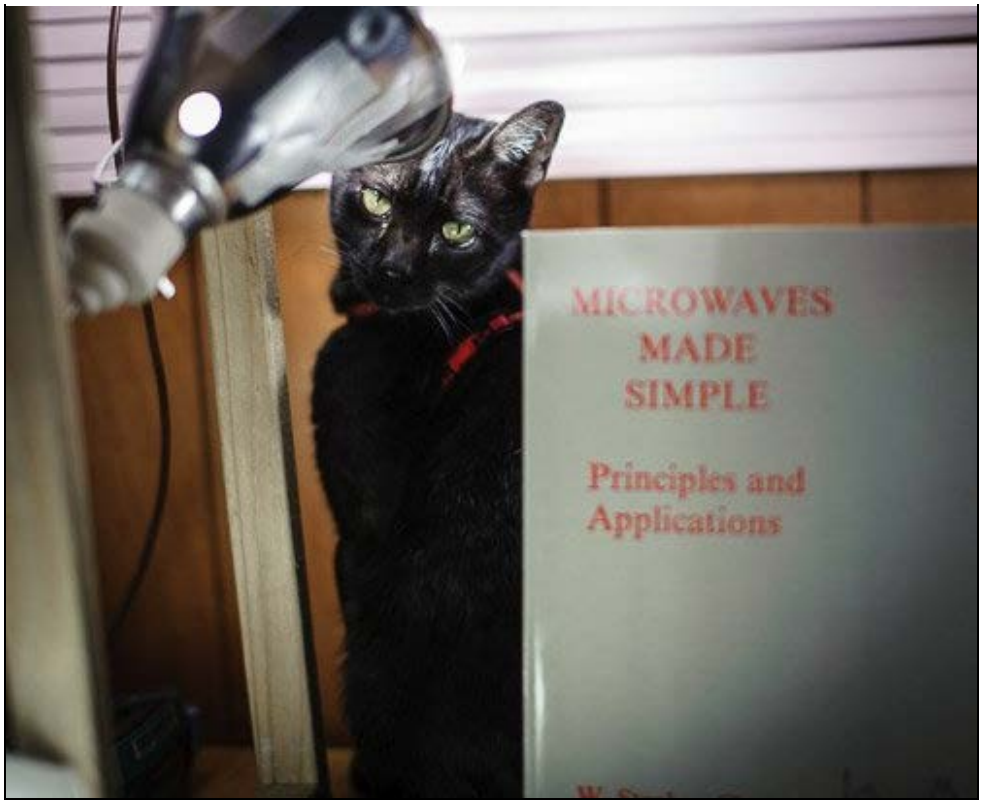
Above — Top of the circuit board. Anti-parallel diodes, 4 MMICS, passives and then even more anti-parallel diodes go before the prescaler chip.

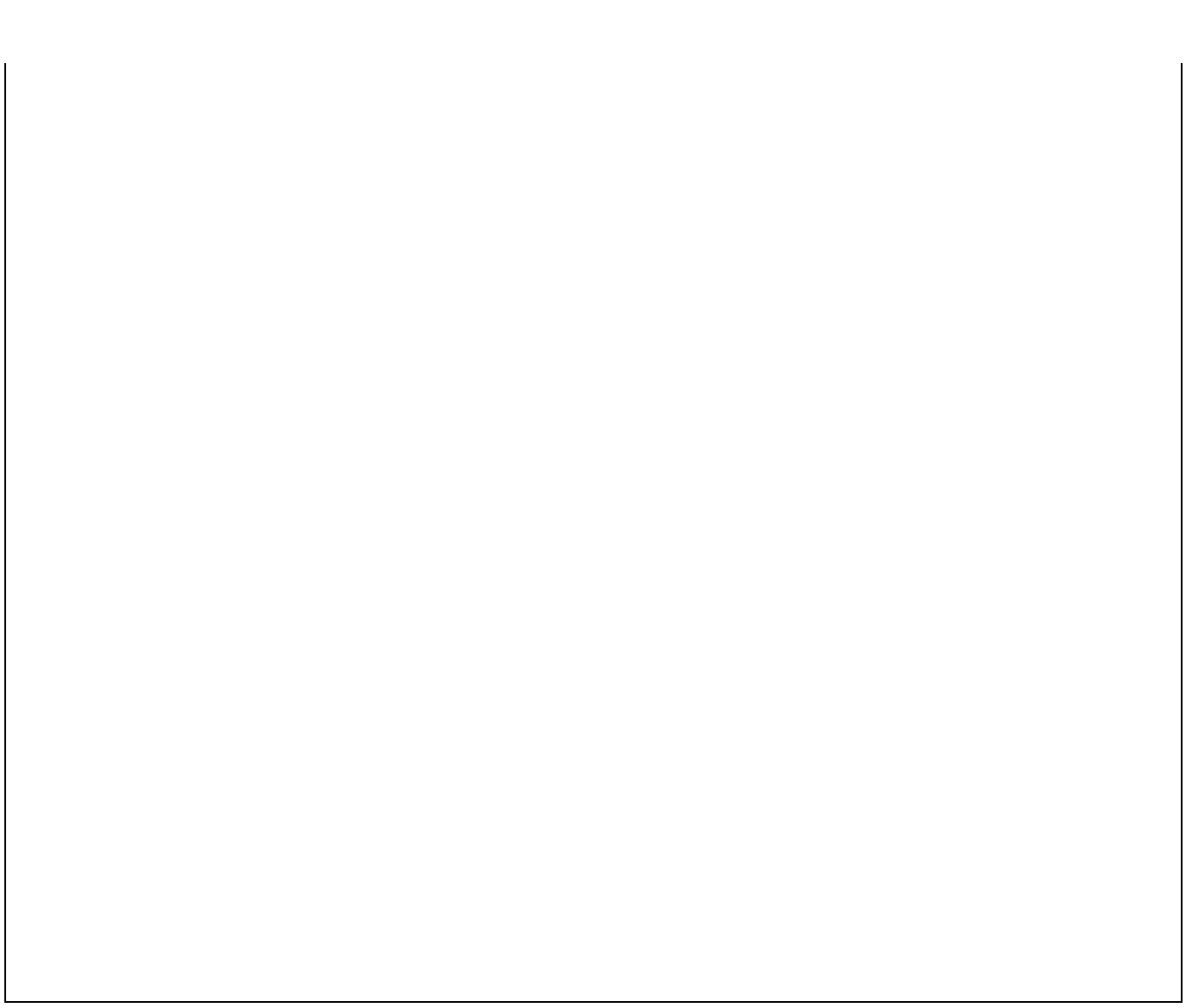


Above — Close up of SMD circuitry.



Above — Rear view of the circuit board showing stiching vias that join the top ground plane to the bottom ground plane along the signal path. My tests showed that it works okay to 3.1 GHz. Click for a picture of the installed board. I'm very happy with my HP counter's boosted bandwidth.


MICROWAVES
MADE
SIMPLE
Principles and
Applications



Funster Line — Transmitter

Menu Item 4 on the HF Embarcadero web page. Click here to go back.

The receiver will go on its own page and hopefully I'll complete and write it up this Fall.

1. Transmitter
2. References

Introduction

I began the Funster transceiver line in 1995 and show the 2014 version [now back to a CW trans-receiver on the 40M band]. Judging by reader feedback, plus analog scratch-homebrew threads in social networks, email lists and print, many of us feel a nostalgic pull towards old-school radio. Me too.

Now at version 8 — Funster shares my joy in making a simple, old-style radio with all discrete parts. In some ways, this radio set is a twisted version of Wes and Roger's Ugly Weekender from QST for August 1981.

1. Transmitter

Summary

Variable power: 128 to 3312 mW @ 13.6 VDC. Switchable 7.04 MHz xtal oscillator or external VFO input port. Solid state transmit/receive [T/R] circuit with PIN diode. External continuous and switched 12 VDC power jacks. Single-ended BJT Class C power amplifier with 67% efficiency.

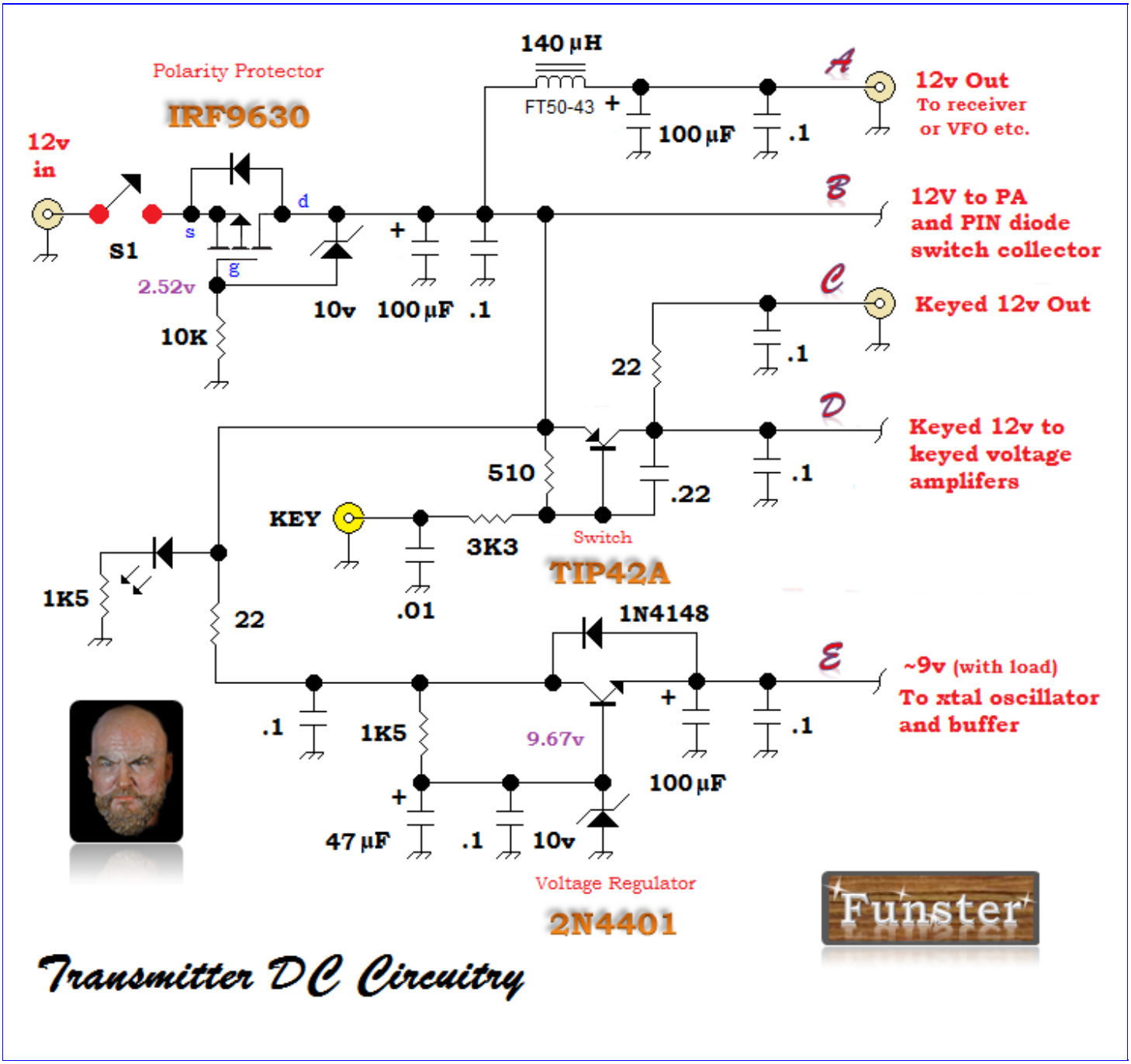
Note: I state 12 VDC in my DC circuits, however, this nominal value might better be termed 12 -13.6 VDC.

Key up power current draw @ 13.6 VDC = 17 mA. Key down current @ 3W RF power with 13.6 VDC = 420 mA.

DC Circuitry

It's fun to march off the signal path and focus on DC circuitry — switching, regulating and filtering the DC voltage feeding our RF or AF stages with finesse. Trans-receivers contain DC, RF, AF and external patch cords that benefit from mindful DC line filtering + selective high-isolation amps in the signal path to prevent RF flowing where it's unwanted. You'll see heaps of decoupling with RF + AF bypass and a few strong

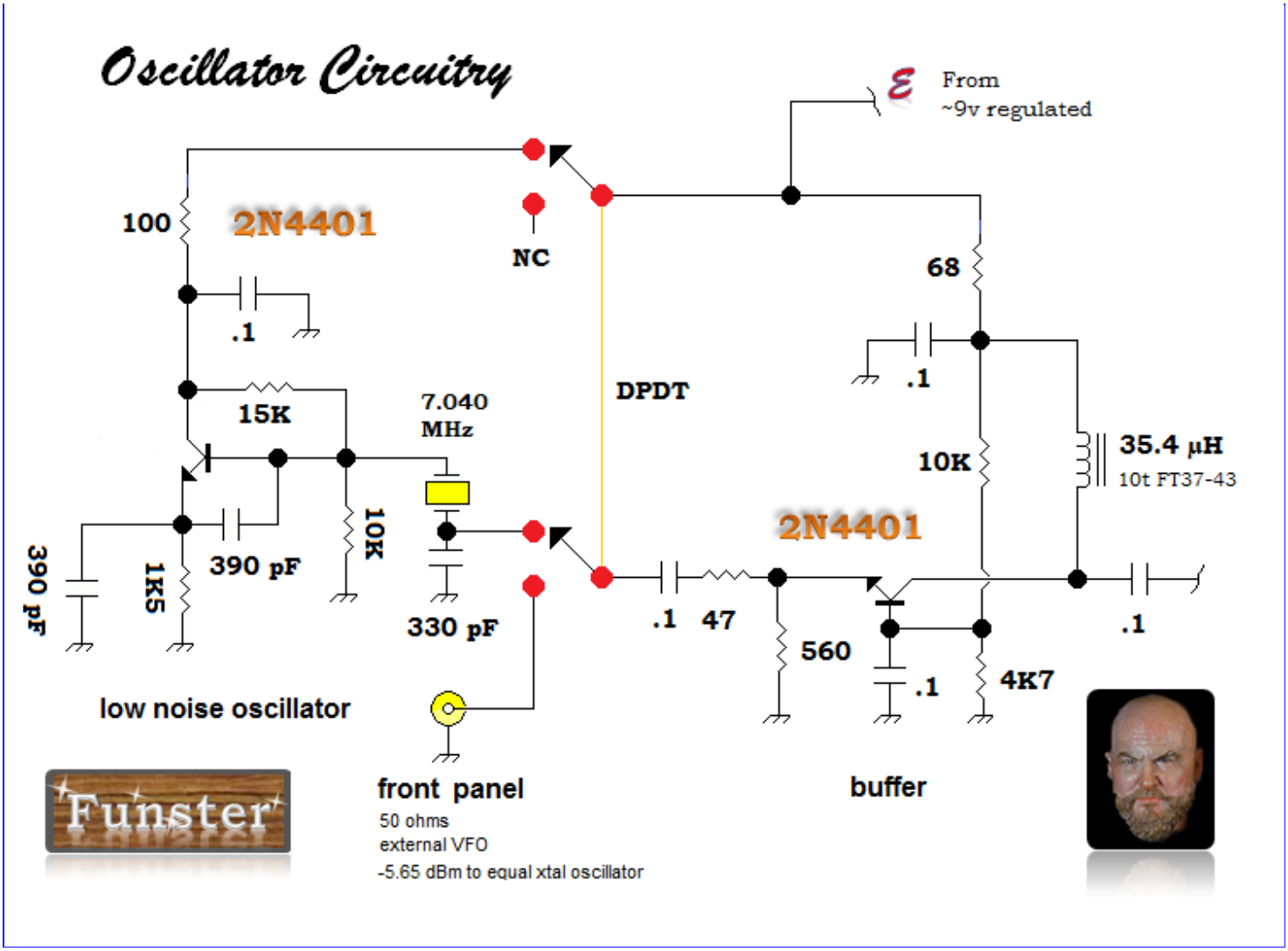
reverse-isolation amplifers in this radio set.



Above — The DC circuit board. Click for the breadboard photo. A reader sent me 2 IRF9630 power MOSFETS: avoiding the voltage drop of the usual single rectifier diode, I employed 1 as a reverse polarity protector. Another reader sent me a long strip of 2N4401 BJTs, so I ran them instead of the equally good 2N3904. Thanks for the parts guys!

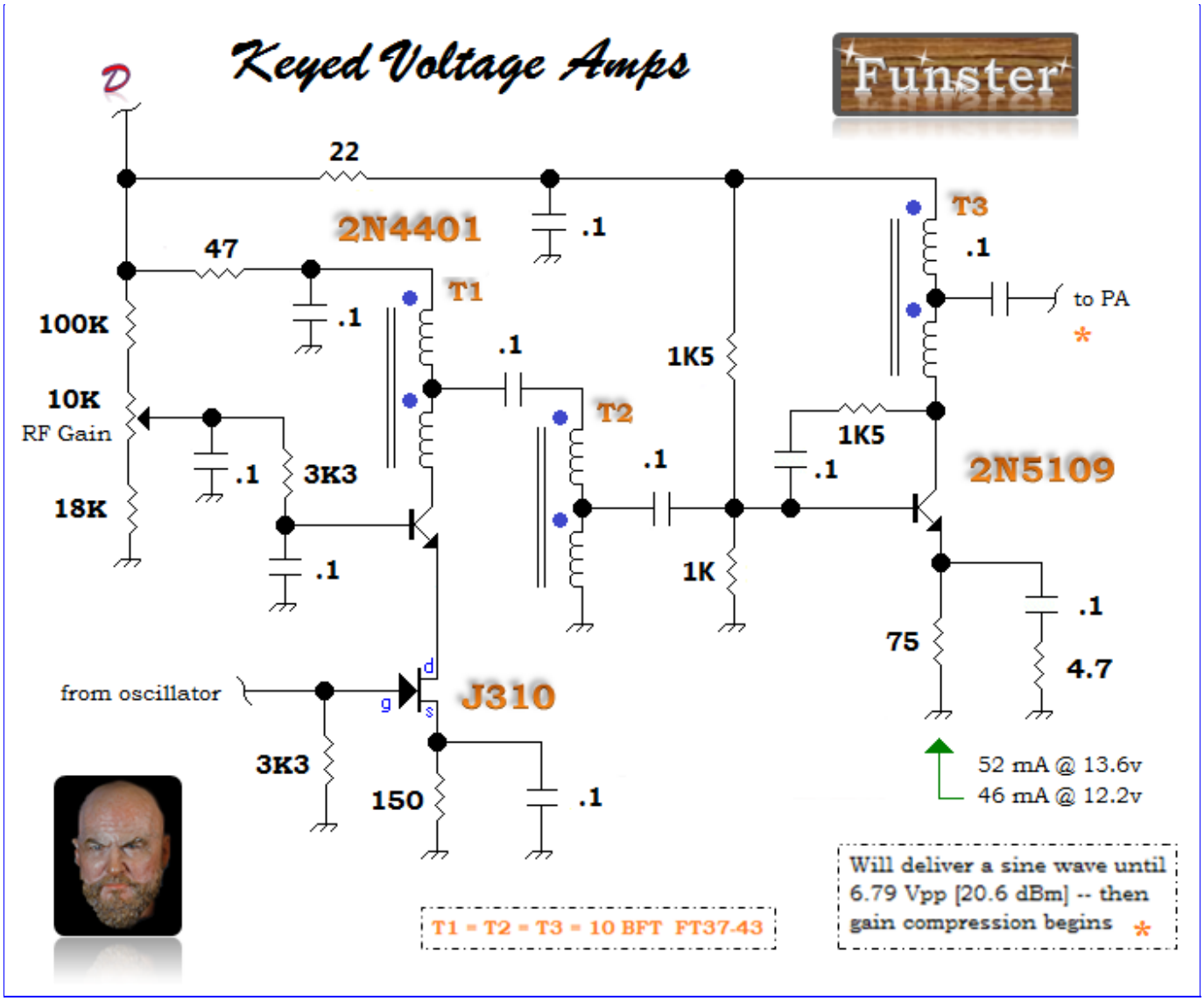
A simple series voltage regulator with a zener diode reference supplies DC to the crystal oscillator and its buffer. The base-emitter voltage drop of the 2N4401 emitter follower lays in series with the load, so load current changes alter the regulator output voltage. Vout = Vzener - VBE : so the regulated voltage is roughly 9 VDC.

Oscillator Circuitry



Above — Oscillator circuitry. I soldered my oscillator circuitry on the DC control board. Click for the breadboard. Click for a close-up of the old NorCal crystal supplied by NT7S — thanks Jason! To equilibrate to the xtal oscillator AC output voltage, the VFO input port should run about -5.65 dBm. I might stick a 6 dB attenuator pad on the front panel switch and use a 7 MHz VFO with 0 dBm output power.

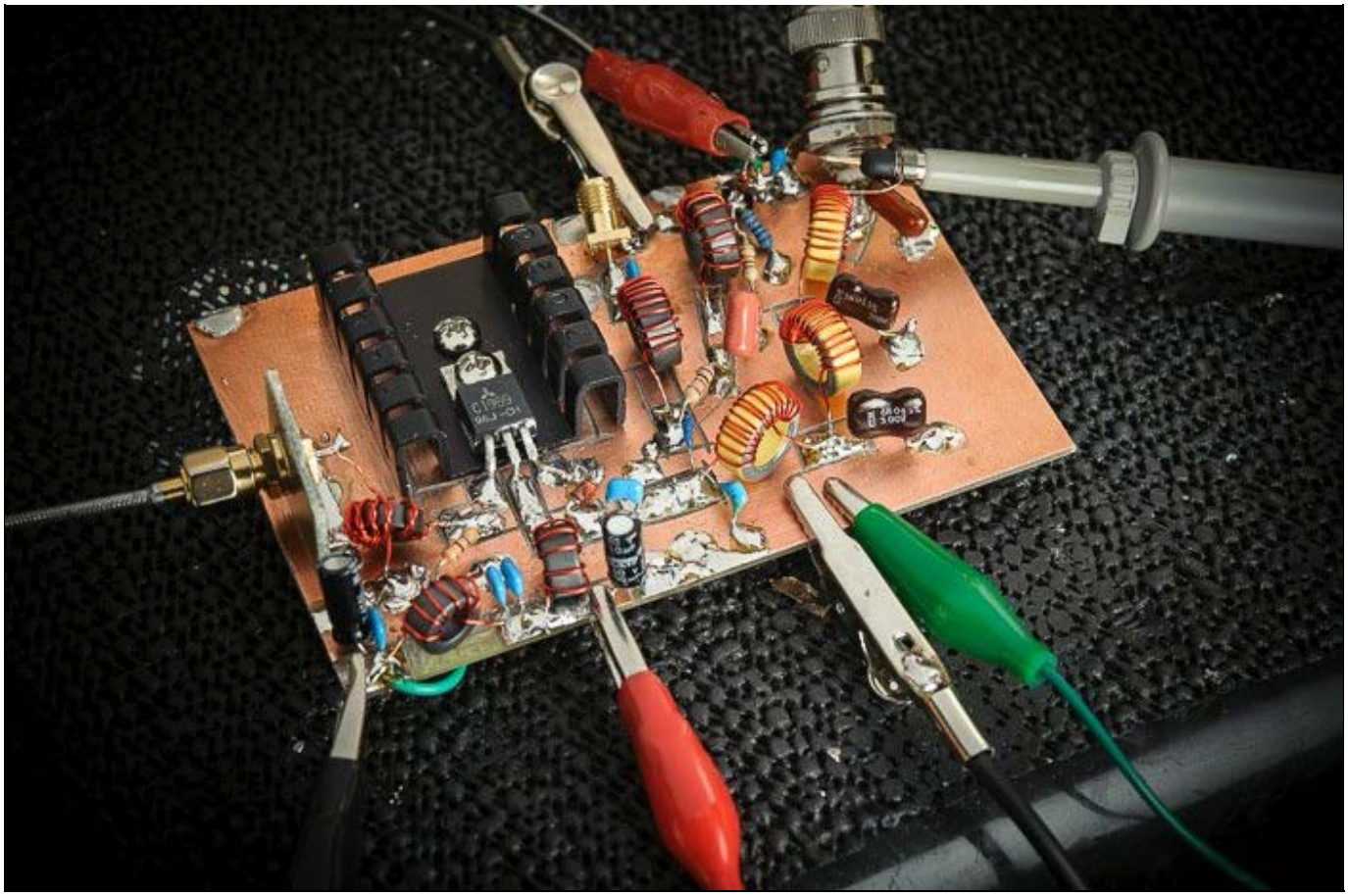
Keyed Voltage Amps



Above — The 2 keyed voltage amplifers. Click for a breadboard photograph. I set the value of the RF gain pot series resistors by placing a 100K pot before and after the 10K gain pot and tweaked them back and forth to find close to 3W maximum and 125 mW minimum RF power with the gain pot set @ CW and CCW respectively. The 100K pots were removed, measured and nearest standard value resistors substituted: 100K and 18K.

By carefully choosing the 75 Ω emitter resistor, the 2N5109 feedback amp output remains a sine wave at the drive levels needed for a PA output at 3 Watts. At 3 Watts PA output, the feedback amp output voltage = 5.91 volts pk-pk [19.4 dBm]. The emitter current shown will cleanly drive a PA for more power if desired. Boost it as needed.

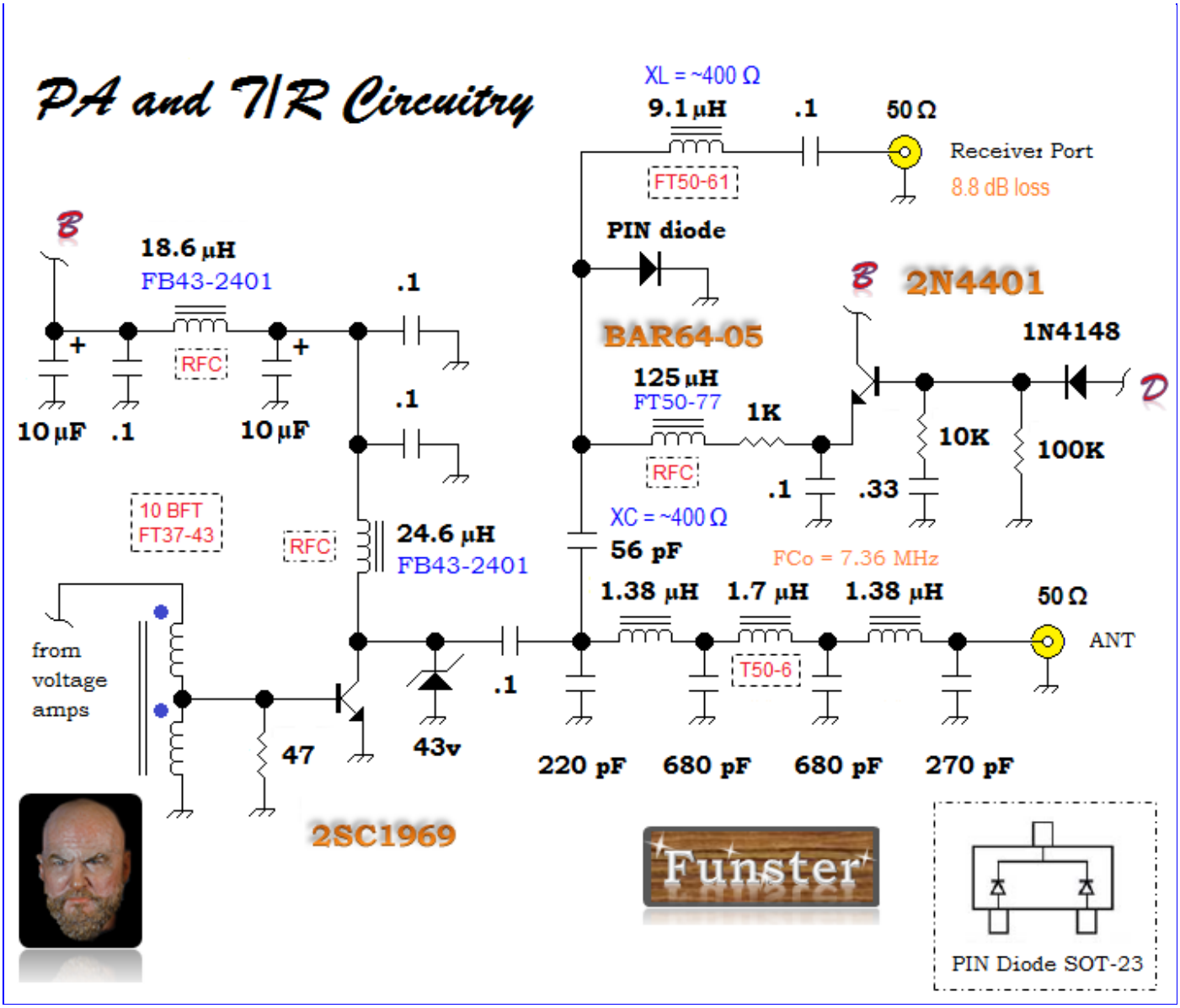
Power Transistor and T/R Switching



Above — Final testing the PA and T/R circuitry before installing it into the Hammond chassis. Click for a bigger photo of my test setup.

Three temporary RF connectors were soldered on this board for bench measurement. I connected the (SMA) receiver port to Channel 1 of my 50 Ω terminated 'scope to measure the RF not getting shunted to ground by the PIN diode. The SMA cable/connector to the far left is a temporary cable connected to the keyed voltage amp board output in the Tx chassis — thus I tested the PA stage with its own xtal oscillator and RF amplifers. The right-sided BNC RF port connects the PA low-pass filter to a 10W, 50 Ω dummy load via a 22 cm 50 Ω patch cable.

Further, I measured collector voltage + collector current with 2 DVMs and also PA output power in my oscillocope (DSO) Channel 2 with a 10X probe employing a tiny ground lead.



Above — PA and T/R circuitry schematic. Click for another breadboard photograph taken before the 2N4401 switch was added. Although we might run a BD139 or other cheap BJT for the final, I chose my favorite HF classic, the 2SC1969. This transistor will survive a high VSWR, however a 43 volt - 500 mW zener diode further protects it from such.

In 1 version with a 1:4 Z collector matching network and no zener diode, I measured 6.25W RF power after tweaking the 100K and 18K RF gain limiting resistors in the preceeding hycas voltage amp to drive the PA harder. At ~3W maximum power in my final experiment, the 2SC1969 barely warms up.

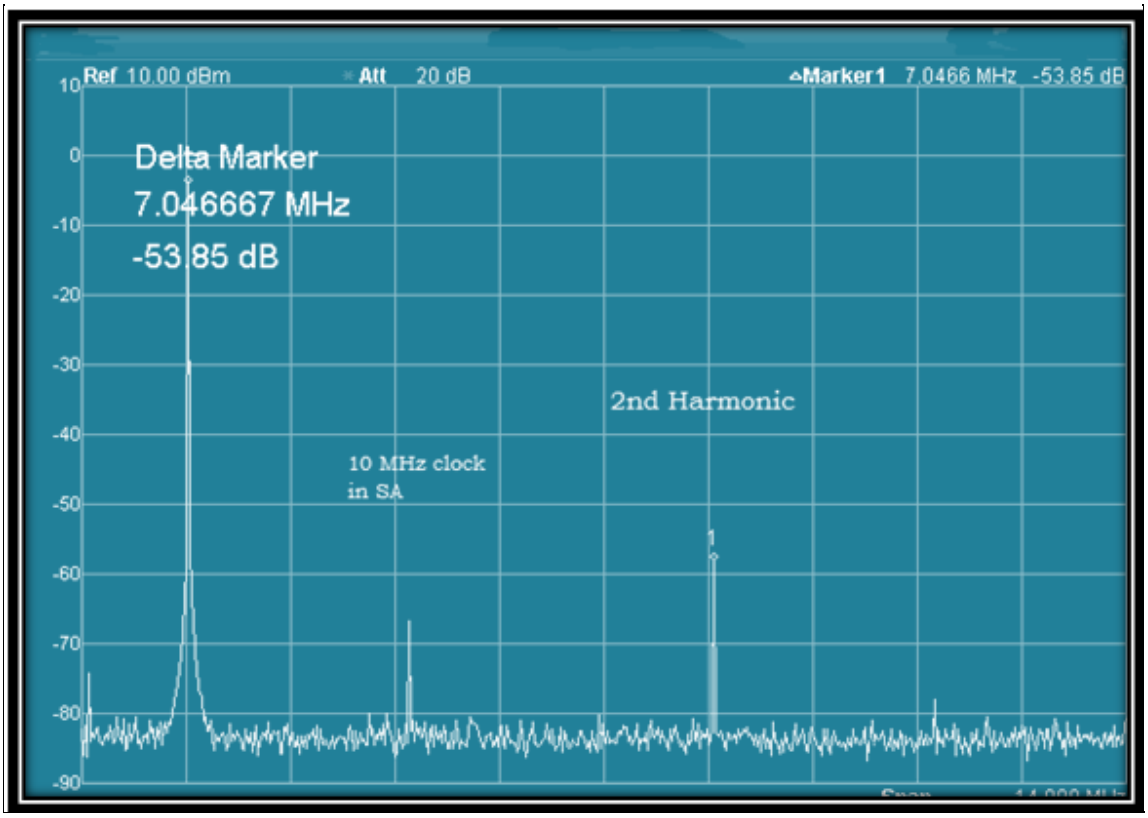
Collector Matching

After a number of experiments, I learned that from tens of mW up to ~ 3.3 Watts RF power, we may omit collector matching and still get reasonable efficiency if you run a collector choke >= 15 µH. Thus, I made the maximum power ~ close to 3W — easy peasy.

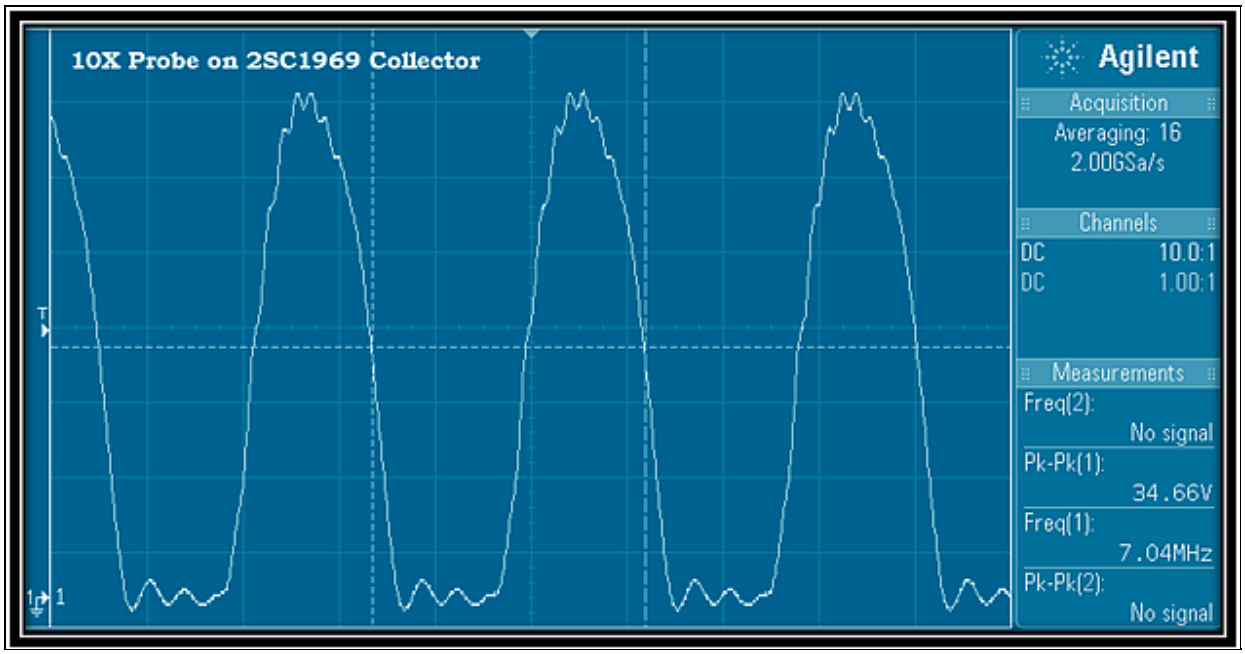
Ask QRP enthusiasts what QRP is all about and they might extol the thrill of exacting the most kilometers per Watt — in keeping with that spirit, I love turning down my Tx power and the Funster front panel RF gain control makes it a breeze.

The PA circuit is standard fodder. I wound the 24.6 µH and 18.6 µH RFCs on FB43-2401 ferrites [a single hole bead that looks like a fat FT37-43]. An FT37-43 or FT50-43 makes a great substitute.

PA Measures



Above — Spectrum analysis of the Tx output. The 10 MHz clock signal from my spectrum analyzer lays between the fundamental and 2nd harmonic. The 2nd harmonic = -53.85 dBc @ 3W power output.



Above — PA collector waveform at 3W output measured with my DSO's 10X probe.

Above — Set your meters on stun! DC measures to calculate PA efficiency [See RF Workbench 6]: DC power = 13.58V x 0.331A = 4.49W. RF power = 3W. Efficiency = RF power / DC power = 3W / 4.49W = 66.8%.

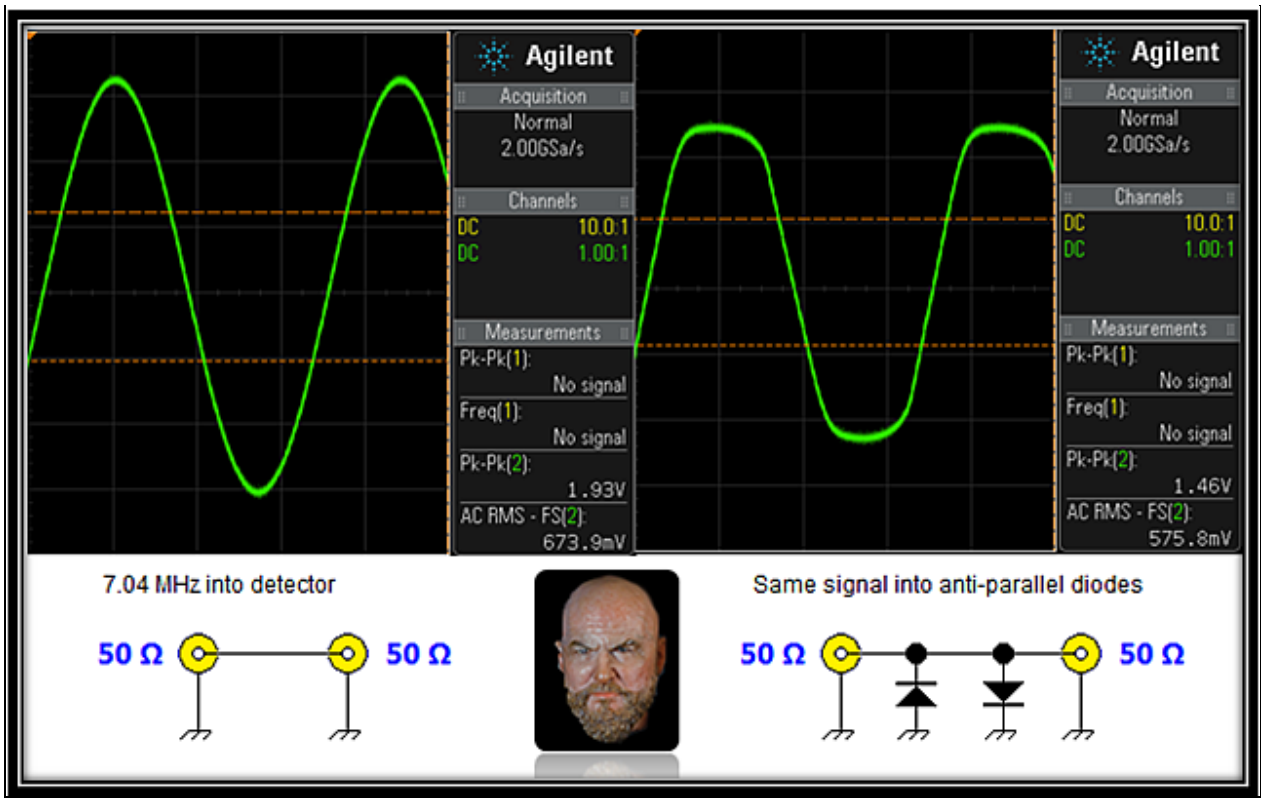
T/R Discussion for Single Band Transceivers

Previous Funsters ran a simple QSK T/R circuit developed by Wes, W7ZOI in the early 1980s: a series X = 500 Ω resonator with anti-parallel diodes between the L and C. The series capacitor presents 500 Ω reactance to the 50 Ω transmit line so that only a small portion of the transmitter power flows to ground via the diodes and even less reaches the receiver input. In short, the 500 Ω capacitive reactance + clamping action of the anti-parallel diodes protect the receiver front end.

For 1 band, this simple circuit sizzles. Two potential concerns:

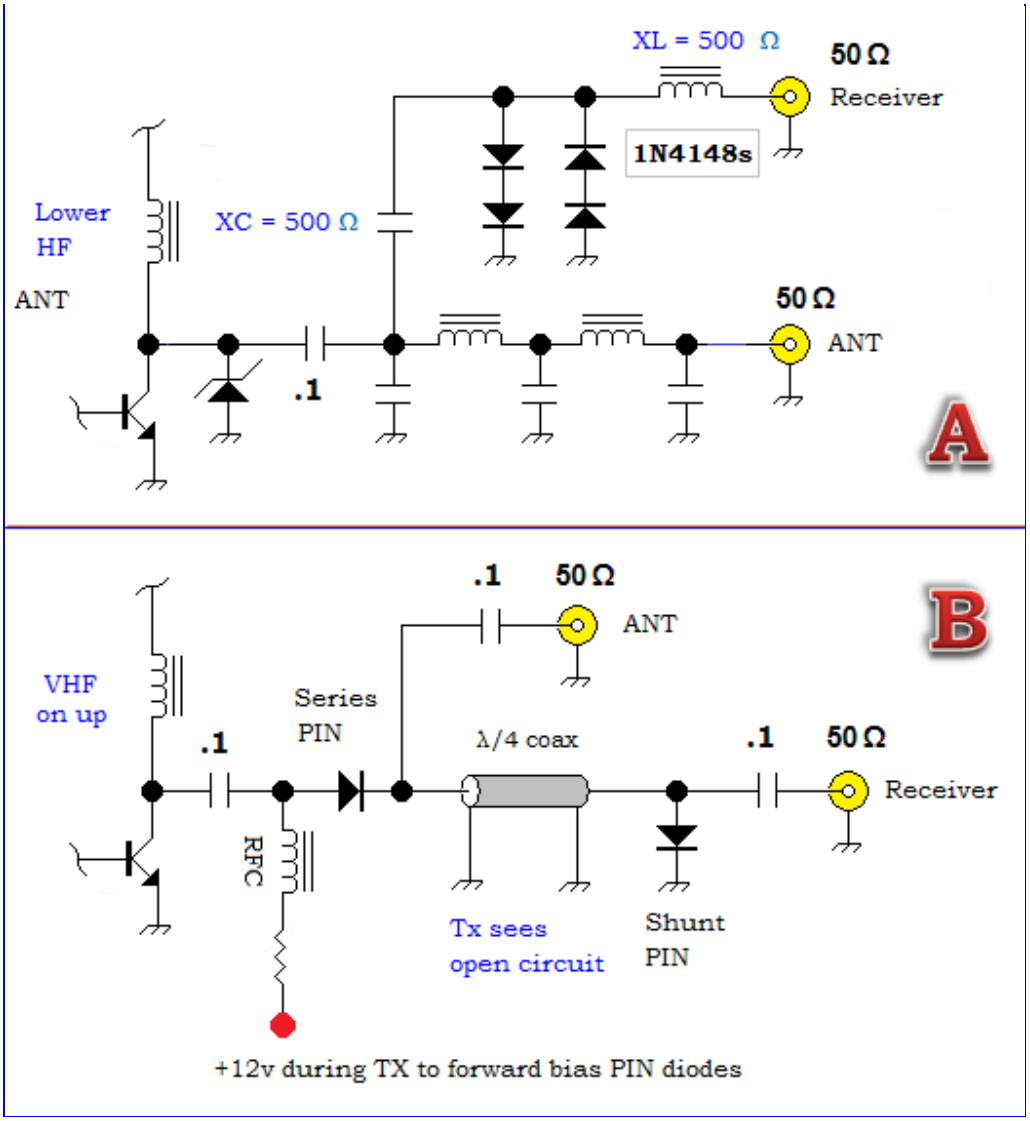
~6-9 dB loss of the receive signal making it better for lower HF, although I've seen builders add a receive preamp and use this T/R system on 12 and 10 meters. Since both band noise and signals get attenuated simultaneously, I've rarely had readability issues on 160-30 meters in my popcorn transceivers.

The anti-parallel diodes may cause IMD. The intercept for the 500 Ohm series reactances with shunt diodes is around +6 or +7 dBm. This might exert problems in a high performance receiver front end, but poses a non-issue with typical NE612 front ends. Many builders, including Wes, reduce IMD potential with 2 or 3 series diodes on each leg. Another solution includes running a PIN diode as the shunt element. That's what I did.



Above — The DSO tracing of an experiment showing the before and after effect of 2 anti-parallel diodes. The voltage across a silicon diode varies with the current going through it. Silicon diodes don't conduct much current until the voltage across them reaches a threshold [typically ~0.6 to 0.7 volts].

In situ, the forward threshold voltage of each diode limits the circuit voltage by clipping off AC voltage above its threshold. Clamping refers to the limiting of voltage in *2* directions using anti-parallel pair(s) of clipping diodes. For example, if a negative-going AC dips more than 1 diode drop below ground, the left hand diode will begin to conduct and clip off any voltage above its threshold. This, of course, describes the "ideal diode" — in real life, factors like switching speed, current-handling and temperature are all in play.



Above — 2 possible T/R schemes. Figure A shows the classic W7ZOI T/R system with 2 diodes per leg. See EMRFD Fig 2.112, for an example with 3 diodes per leg. For 1 band, it's often our go-to circuit.

Figure B shows a scheme well suited for VHF on up: Both PIN diodes get forward biased during transmit mode, however, a 1/4 wave transmission line lays between the series and shunt PIN diode. 1/4 wave transmission lines exert magic [actually it's pure physics that may be verified with a Smith chart]. When the shunt PIN diode gets biased *on* and grounds the receiver port end of the transmission line, the other end looks like an open circuit and thus power flows to the antenna port and not into the forward biased shunt PIN diode to ground. I hope to apply this system in future VHF and UHF projects and sadly, it's impracticable at HF.

Funster Version 8 T/R

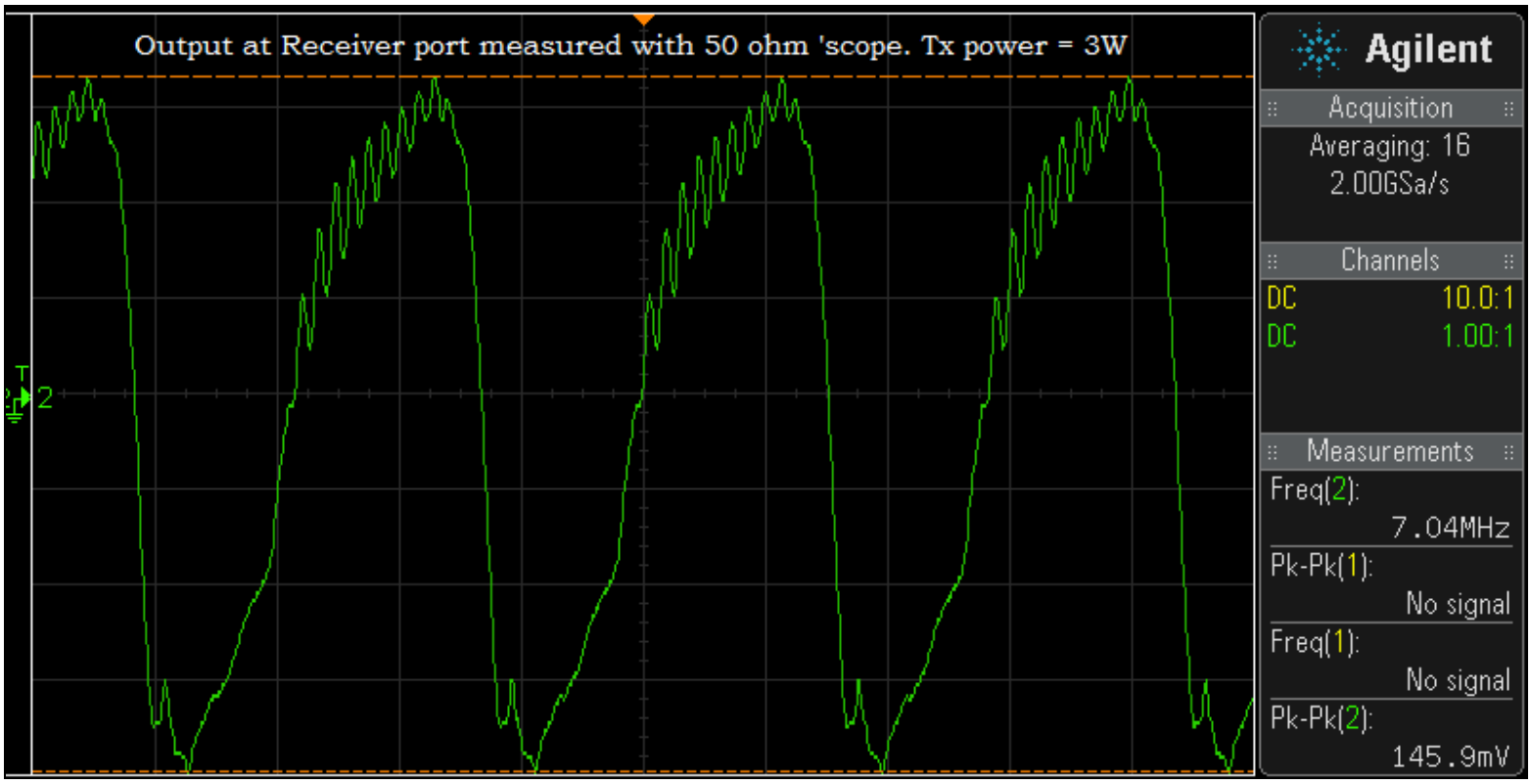
As it happens, I seek experience with PIN diodes and therefore shunted a PIN diode instead of running anti-parallel diodes in the classic W7ZOI T/R scheme. This meant adding a transistor switch to quickly bias on the PIN diode @ *key down* and run a little hang time @ *key up*.

Please refer to the PA-T/R schematic. During key down, keyed 12 volts saturates the 2N4401 switch and 11 ma current biases *on* the PIN diode. At key up, the BJT switch's base voltage is held on by the 10K plus series 0.33 µF capacitor just long enough to allow the stored energy in the large PA collector choke to collapse — protecting the receiver.

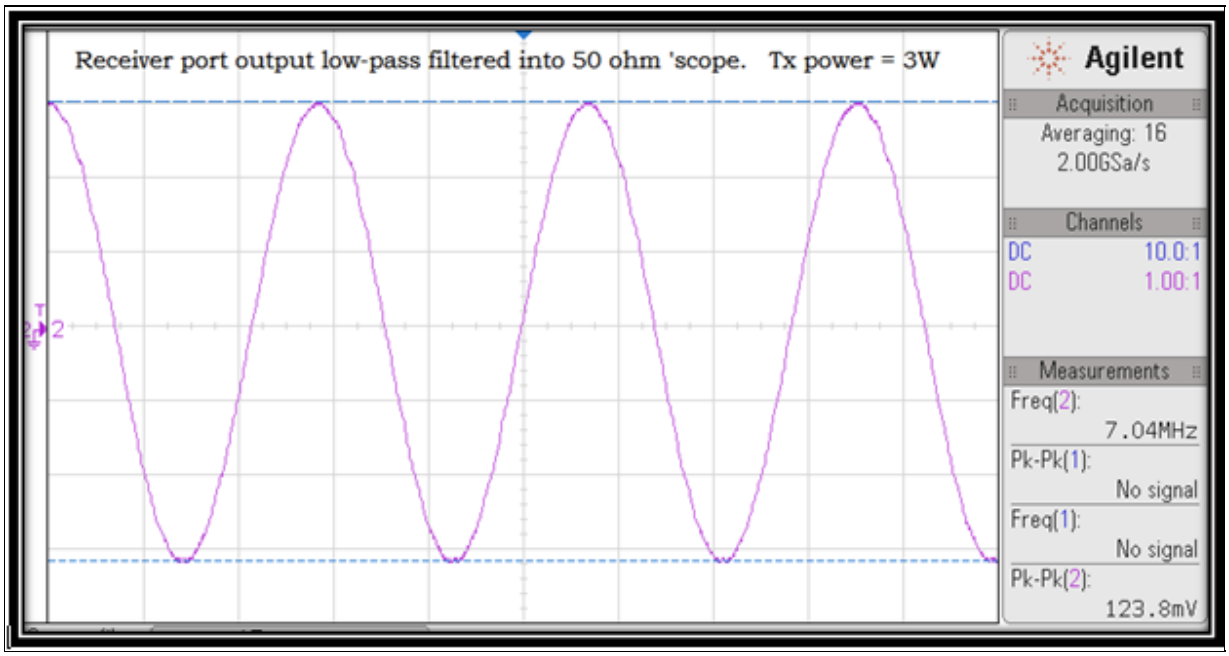
Feel free to experiment with the timing capacitor value. Previously, Wes, W7ZOI suggested connecting the T/R receiver port to a 50 Ω teminated 'scope for general measurement and also for clocking up RC time constants in DC switching circuits. With a 2 channel 'scope you can monitor the receiver port switch function in 1 channel while watching the PA collector AC voltage in the other. Now that's experimenting! Click for a sample from my QRP workbench.

I chose an XC / XL of 400 Ω for the series L-C network. The 9.1 µH coil shouldn't be wound on a 43 material ferrite [the Qu is just too low]: a bigger powdered iron toroid like size 68, or even a 61 ferrite material toroid works okay from my experiments [Q's as high as 162 were measured on a FT50-61 @ 1 MHz in 2013]. Although this coil tunes coarsely, try tweaking the 9.1 µH inductor with thumb and forefinger during actual signal testing in a receiver.

I wound the 125 µH choke connected to the pin diode cathode on a 77 material ferrite toroid to get it done with just a few turns and avoid a self resonant frequency close to 7 MHz.



Above — 50 Ω terminated DSO measurement of the receiver port @ 3W transmitter power. Theoretically all power should shunt to ground through the PIN diode, however, the diode exhibits a little resistance and the series L-C circuit contains reactance, so isolation is never perfect.



Above — A low-pass filtered measure of the receiver port while in transmit mode. The beauty of a 50 Ω measurement environment shines: I placed a 7.36 MHz-bench module low-pass filter between the receive port and my DSO to quantify the leaked power to the receiver port as -14.17 dBm when the transmitter delivers 3W. Safe for my receiver.

My sincere thanks to those who offered advice and feedback during the build of this transmitter.

Above — Funster in the field.

Above — Rear view of the Funster Tx.

2. References

T/R References

Perfecting a QSK System. QEX for Mar/Apr 2006. Markus, VE7CA. Click for pdf.

Electronic Switching and How it Works, QST for September 1984. Doug, W1FB

7 MHz receiver -- discussion begins on page 215 -- On page 214 in Fig 1 see the 33 pF input cap plus receiver protecting anti-parallel diodes. Also view the 555 timer and other T/R control circuitry in Figure 2 on page 215.

This receiver realizes the first-ever printed version of the W7ZOI T/R system described earlier. Solid State Design for the Radio Amateur. ARRL,1986 [out of print] by Wes, W7ZOI and Doug W1FB.

Transmit - Receive Antenna Switching. EMRFD. First published by the ARRL in 2003. Page 6.68.

Schematic for the 30-W, 7 MHz power amplifier. Figure 2.112 [page 2-42] EMRFD. Design by Wes, W7ZOI

VHF FM — Supplement 1

This page supplements VHF FM

1. 10.7 MHz IF Strip - with example broadcast band FM receiver
2. NOAA Receiver Experiments @ 162.55 MHz

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1. 10.7 MHz IF Strip

I'm inexperienced with making FM receivers and lack confidence. I began with a utility module to allow basic assessment of the circuit gain required for different detectors, ceramic filter bandwidths, and more so — to get something working to boost my confidence.

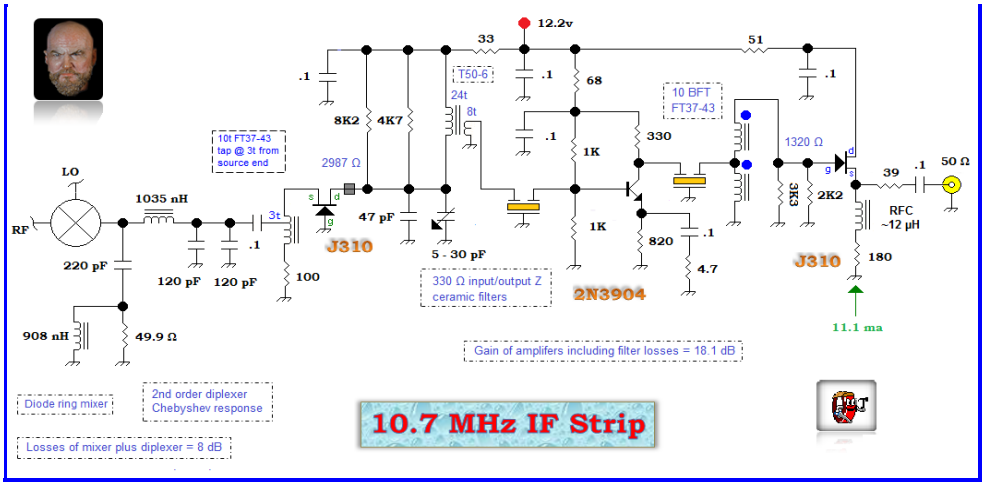
The FM receiver 10.7 MHz IF standard began soon after World War 2 and 10.7 MHz still makes sense for the primitive radio builder — surplus IF filters still seem available in a variety of bandwidths. Although vexxing, if we suffer from receiver images at 10.7 MHz, we can always filter the input better, or jump up the IF and make a crystal filter for say 20 MHz. Above 20 MHz good, inexpensive, fundamental-mode crystals get hard to find.

Mixer

I chose a diode ring mixer; again. After trying other mixers such as balanced FETs, or an NE612 , I kept coming back to the diode ring. Partly I'm biased and partly it's from from measures like this one. Although the oscillators weren't perfectly set for a 10.7 MHz IF, the LO and RF signals are both > 40 dB down from the desired IF — love this.

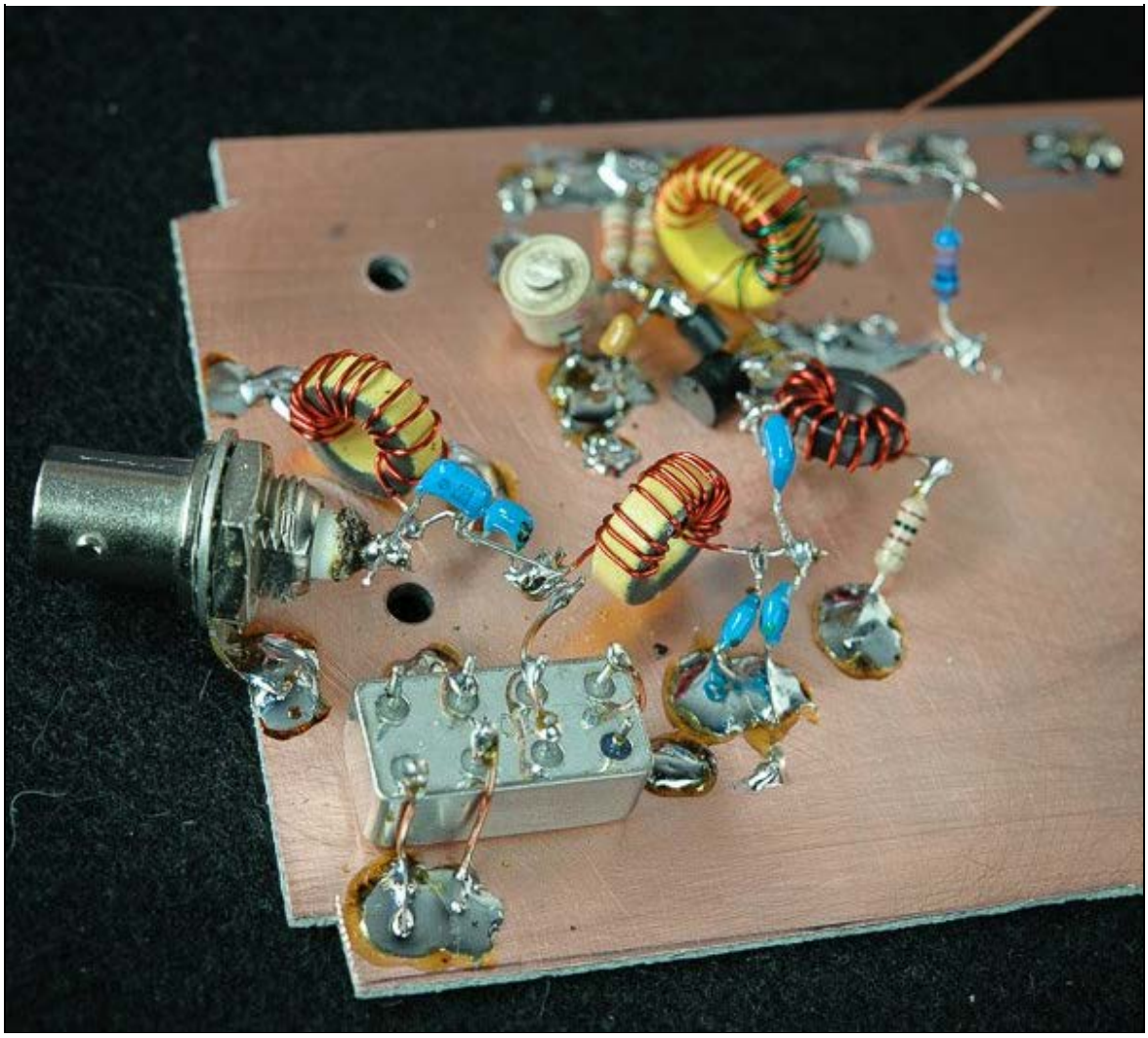
Since I designed a 50 Ω bench module, the IF output contains a source follower to drive the IF port. In an actual "keeper" receiver, you probably wouldn't bother since you might just directly drive the detector, or perhaps another 330 Ω input/output IF amp before the detector.

Although I humbly build simple/primitive circuits, my design goal is never for a *low parts count* — each part serves its purpose. A low parts count ideology plus minimalism exploits the frugal and primitive homebrew nature of our brother hobbyists, however, often gets out of hand. Some builders just cite a low parts count *design goal* to rationalize a poor design. Worse yet, inexperienced minimalists often exclude vital parts such as DC supply decoupling and bypassing, temperature stablizing bias networks, or adequate band-pass filtering. To each his own. Please discern all you see or read in homebrew radio; we're builders, not lemmings.



Above — Click on the IF strip schematic to view it. Most all the stages were discussed on VHF-FM. My net module gain = 10.1 dB. The 2 ceramic IF filters = 230 KHz Muratas with a measured average loss of 6.15 dB each.

Above — The boxed up circuit contains 3 SMA jacks. Click for a photo of the completed breadboard.



Above — A close up of the mixer, diplexer and JFET amp, although the SBL-1 mixer is disconnected. This board contains a temporary BNC jack to allow testing of the non-mixer parts such as JFET gain, diplexer loss, return loss, to confirm the tuning cap choice and also to look for any UHF oscillations eminating from the FET, a UHF part. Click to zoom in on the common gate amplifer with its gate lead soldered as close to the plastic case as possible.

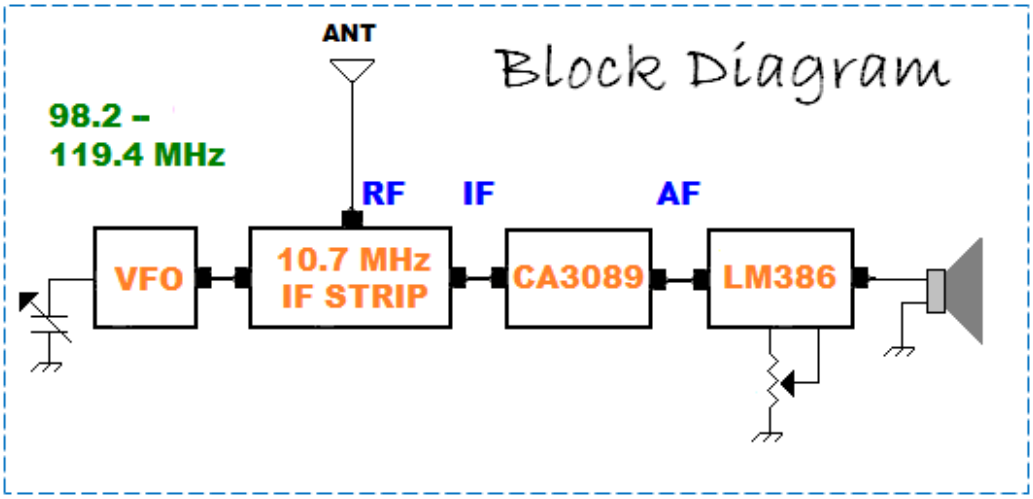
Detector

I tried to homebrew my first IF amp +/- limiter and detectors , but felt unhappy with their performance, failures, circuit complexity and/or noise issues. Click click or click for 3 raw, miserable design examples from Winter 2014. Discouraged; after 2 months of lamenting, I opted for an IC detector subsystem. 3 relic part choices might include the CA3089, SA604A or LA1150.

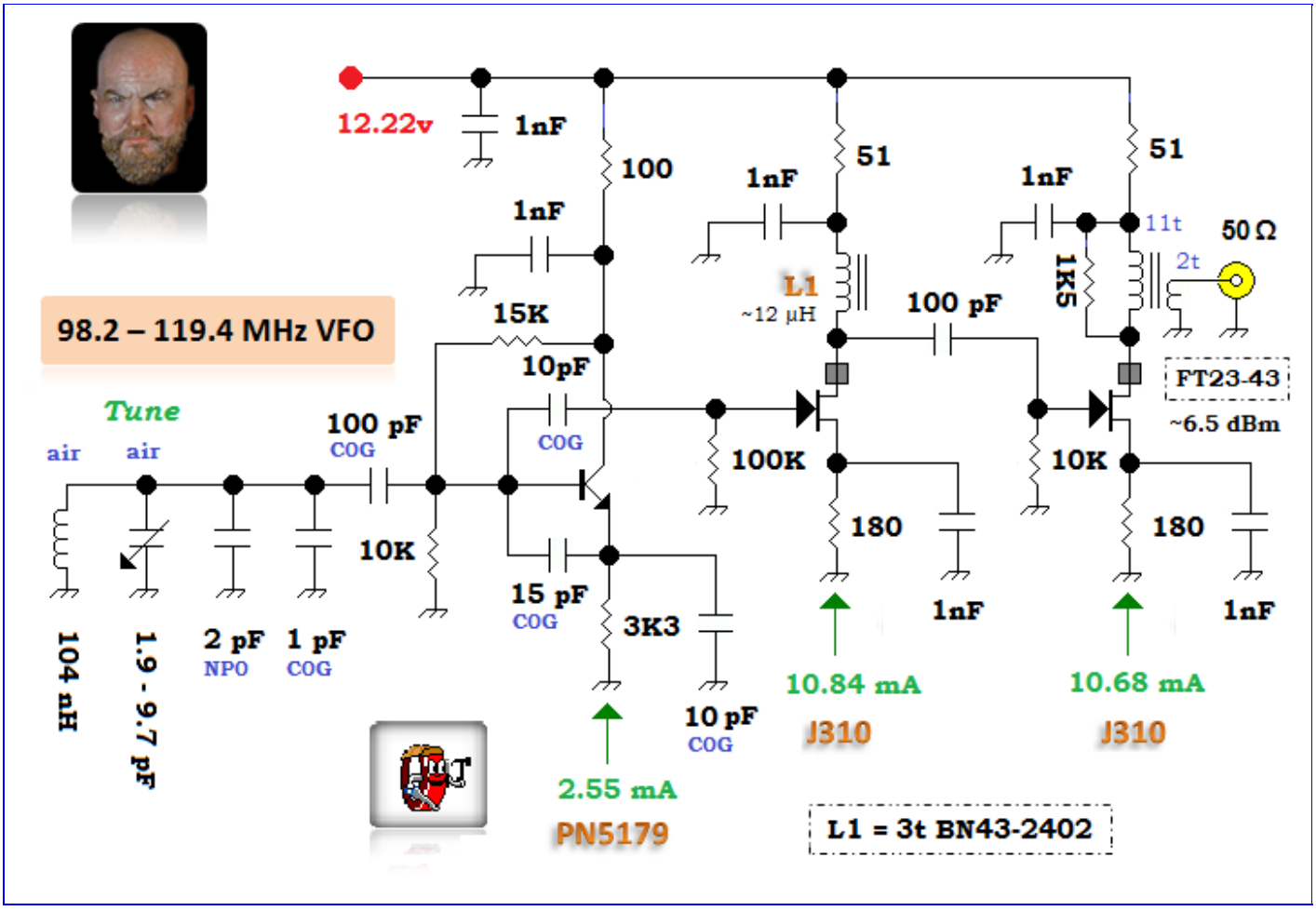
In my case, the CA3089 suited me best and offers a lot: a double-balanced quadrature FM detector with limiter + IF ampifiers; squelch, AFC, AGC, AF preamp and the potential for high-fidelity FM reception. All that's required is an outboard detector resonator — easy to build with a powered iron toroid plus a fixed + trimmer cap.
Price + shipping varies widely, but last year, I got 5 shipped to me from Europe for $7.25 USD and these will make 5 receivers — not bad at all. What hobby isn't expensive?

I've even read that some builders used the CA3089 for an S-meter amplifer, although its logarithmic compliance is far from stellar. You'll need to download the CA3089 datasheet.

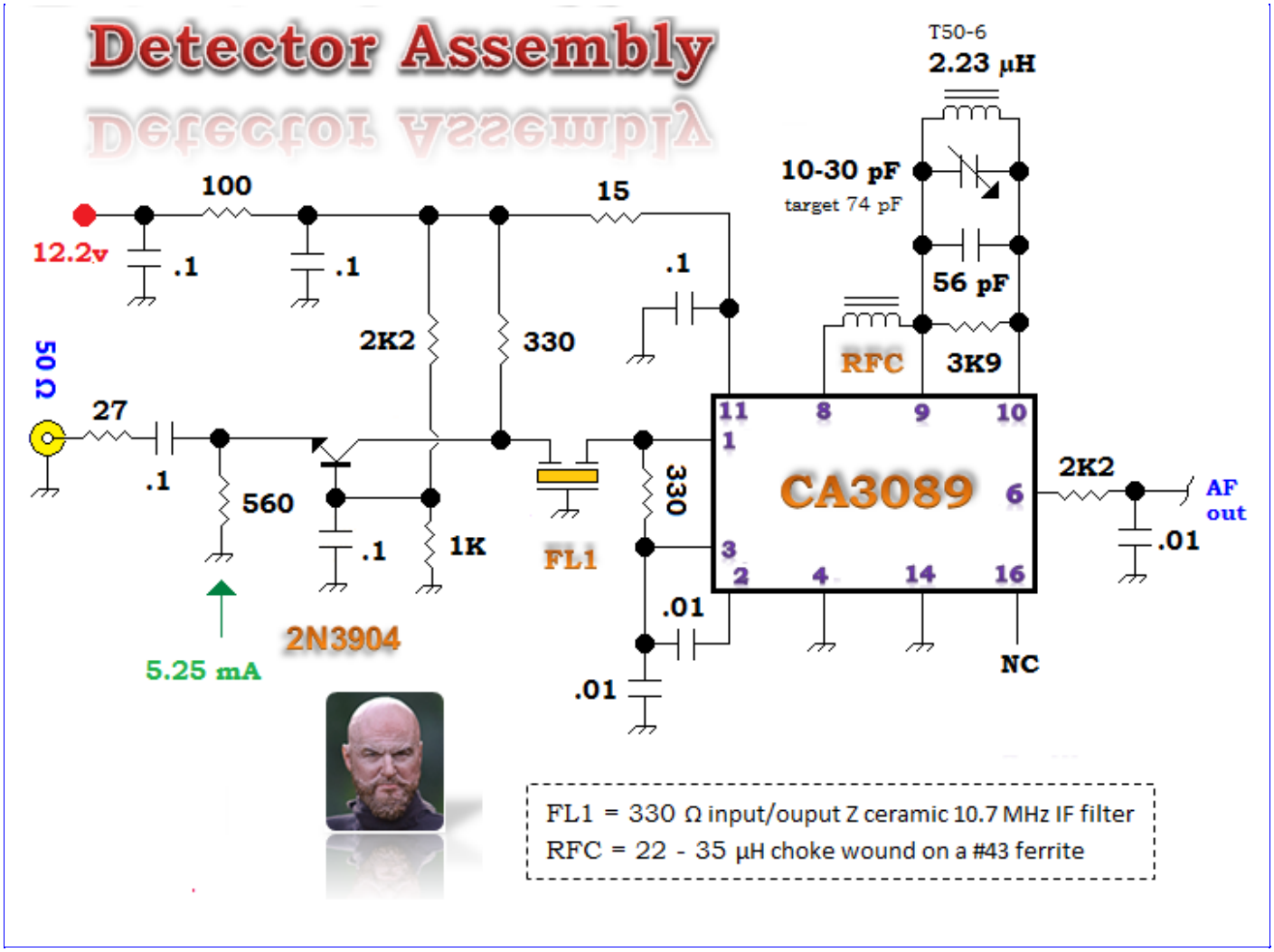
Let's build a receiver:



Above — The block diagram. The antenna = 1 meter piece of wire soldered to a SMA jack that was threaded into a brass sheet to fashion a crude ground plane. No LNA or input band-pass filtering was employed.



Above — The temperature compensated, free running VFO. Normally we should run a VCO-PLL, digital contraption, or a VFO/VCO with ALC feedback, but this thing worked fine to learn on.



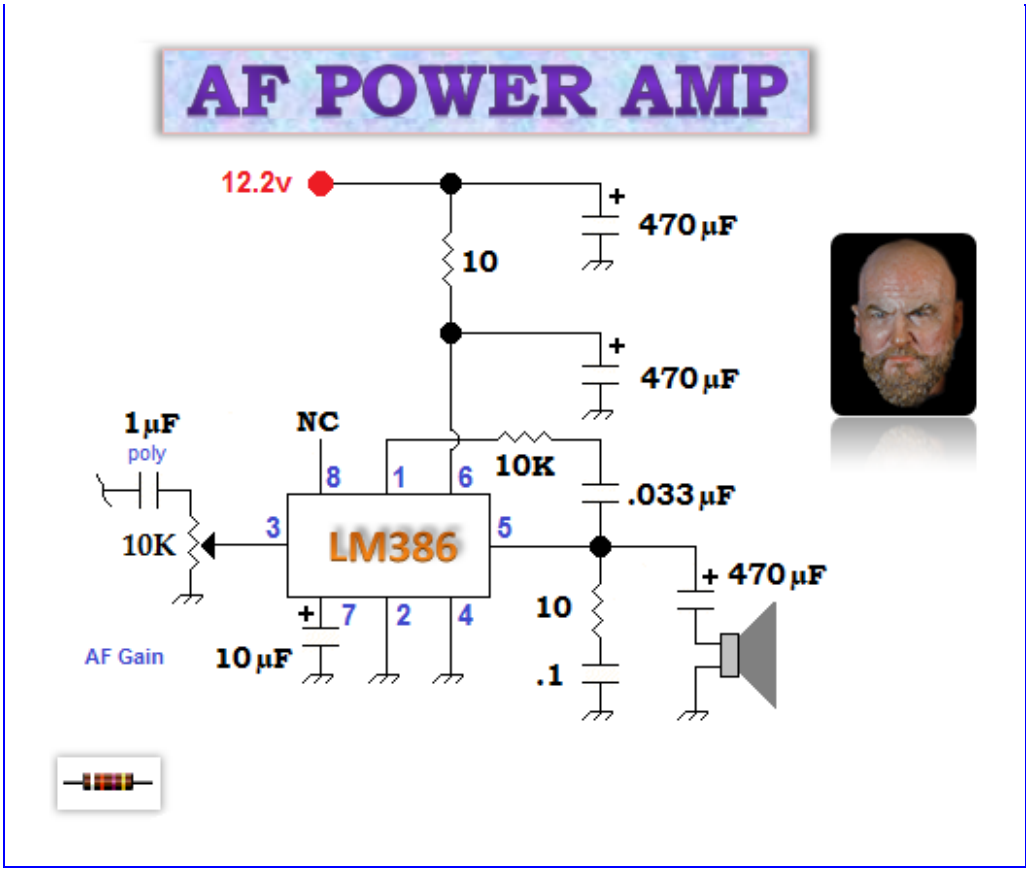
Above — The detector assembly. To match to the 50 Ω IF strip port, I employed a common base amp with a 330 Ω collector resistor driving a final 230 KHz wide "tail end or mop up" filter. In many designs you could simply drop this amp as the gain isn't really needed. On the other hand, the CA3089 contains a limiter and it won't hurt at all.

To make the detector coil, a small square island was carved on the copper clad board. The island was bisected so that pin 9 soldered to 1/2 the island and pin 10 to the other half. The 3K9 R, 56 pF C and a small, high Q SMT trimmer capacitor bridged the 2 islands. Finally, a 2.23 µH, T50-6 toroidal inductor was soldered across the 2 islands.

Tune a station and peak the trimmer cap for the most crisp sounding audio — it's set and forget. When driven with enough signal for full quieting, the CA3089 delivers very low distortion — the FM receiver in my truck sounds worse.

An RFC provides a DC short across pin 8 to 9 and keep ~5.5 VDC bias on the pin 9 transistor base. The minimum inductance should be 10X the detector coil. I wound my RFC with 6 turns on a BN43-2402 [35 µH] , but any old #43 ferrite wound choke >= 22 µH should work. For my chokes, I prefer a low # of turns on a binocular core to avoid any SRF issues.

Since the CA3089 contains a powerful audo preamplifer, only an AF power amplifer with low gain is required. I opted with the old LM386, but in future, better, "keeper rigs" I'll add an op-amp tone control stack driving a pair of power followers for good measure.



Above — Final audio power amp: an LM386 set for a voltage gain of 20 with some low-pass filtering and stout DC line filtering.

Click for a photo of the detector and LM386 assembly. Click for a better view of the outboard resonator and nearby RFC.

Outcomes

After connecting the VFO, antenna, IF strip, detector assemby, speaker and powering them up, loud FM radio blared through the loudspeaker. I tuned every local FM station with full quieting and felt surprised that despite no band-pass filtering, plus no RF preamp, it worked and sounded so well.

Further experiments showed that inserting as much as 15 dB attenuation between the IF stip and the detector assembly still kept full quieting on all the stronger stations. In some cities with multiple, strong FM stations, input band-pass filters will be needed. I'll cover that in a later installment.

I connected this VFO to the IF strip and tuned in my local NOAA weather station at 162.55 MHz. Click for a photo of the VFO built on a piece of bent copper sheet metal.

Future experiments may include changing ceramic filter bandwidth, trying different mixers, applying varactor-tuned input band-pass filters for strong FM environments and for FM Dxing. Finally, I'll design a low-noise preamp with a common gate JFET and/or a dual-gate MOSFET. Utimately, I hope to make a dedicated NOAA weather receiver, a broadcast FM Dx receiver and then some narrow band FM jobs for 2M or 6M. The future looks good.

Audio sample one : a few local stations [mostly recorded before I added the negative feedback on the LM386]
Audio sample two: a weak signal [our provincial Highway Report] that doesn't reach full quieting.

Thanks.

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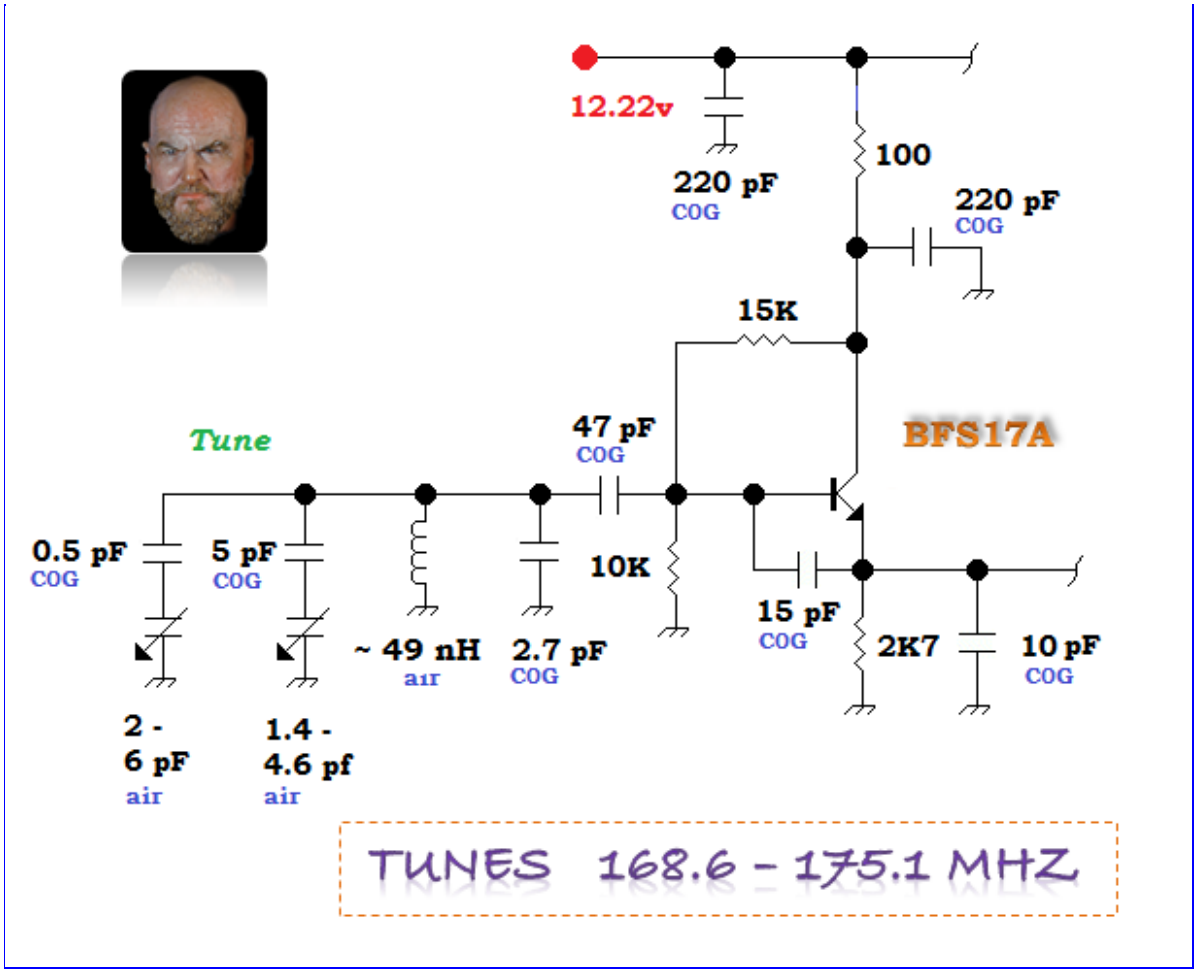
2. NOAA Receiver Experiments @ 162.55 MHz

I document some *raw* experiments and hopefully 1 day will make a good receiver for my local NOAA FM broadcast.

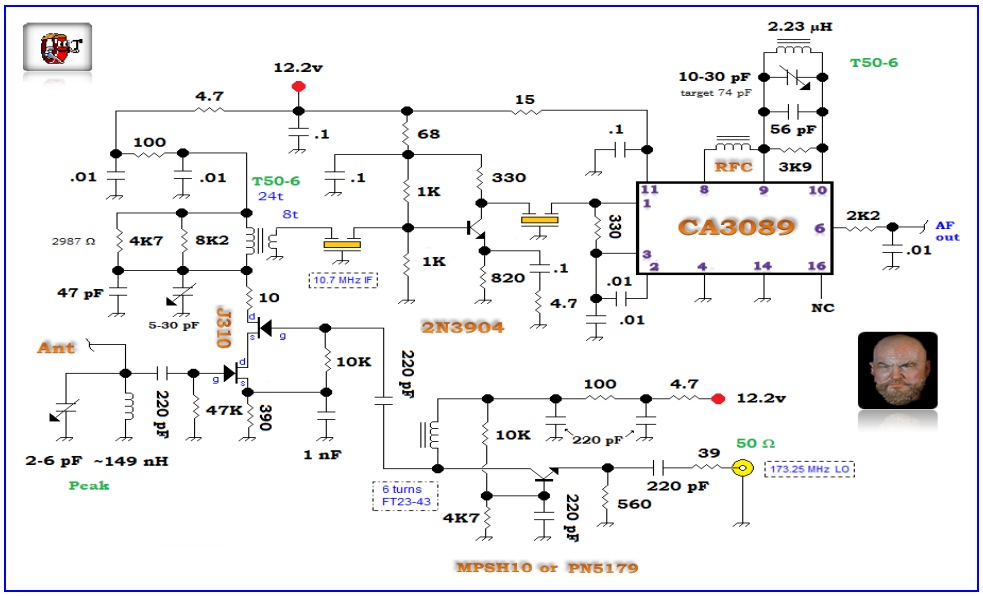


Above — I built a VFO for 173.25 MHz with fine tuning for my local, NOAA weather report receiver @ 162.55 MHz. The fine tuning 2- 6 pF variable capacitor in series with an ATX 0.5 pF chip cap resolves 10's of kilohertz to permit tuning narrow band signals. You'll especially need fine resolution tuning when applying ceramic IF filters <= 150 KHz.

Above — The 2 VFO air variable tuning capacitors. Click for the circuit breadboard with a tuned, common emitter buffer, that I'd rather not show [because it sucks]. I ran the VFO into a bench module to boost the power up between 5 and 10 dBm in my experiments.

Since @ VHF, my VFO went on 2-sided FR4 board with several wire vias joining the 2 surfaces. We normally avoid double sided boards for VFOs since doing so makes a low Q, potentially moisture absorbant, undefined temperature coefficient "capacitor" that may worsen temperature

drift. For bench learning and in particular, FM reception, temperature stability tested adequate and I can listen for hours without retuning.



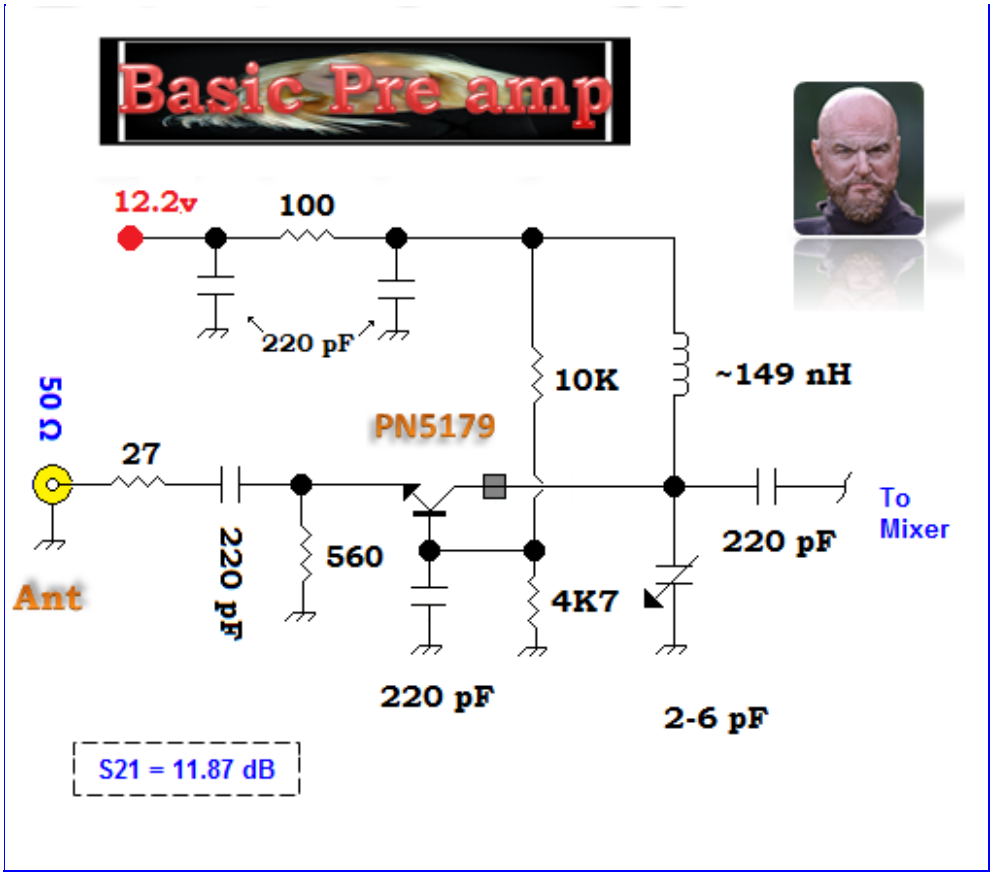
Above — The entire receiver with a cascode mixer from EMRFD. In experimenter's mode; I tried out many ideas. For example: different bandwidth IF filters [settled on 230 KHz] , 1 versus 2 ceramic filters, or eliminating the 2N3904 amp [2 better drove the CA3089]. One salient learning — IF filters may possess a wide variation in center frequency and thus if you have a sweeper, matching them might be useful. In this receiver topology, you don't need to match 230 KHz or 280 KHz ceramic filters; the mismatch became noticeable with a <= 150 KHz 3 dB filter bandwidth.

I boosted the VFO signal with a common base amp — I varied the power into this LO amplifier by applying different strength attenuator pads. Between 1.5 to 2 Vpp on the mixer gate gave a reasonable compromise of conversion gain, noise and easy-to-get VFO power. From EMRFD, maximum conversion gain occurs when driven with 5 Vpp.

This unbalanced mixer worked fine, however, without a metal case, I heard a little hum from my DC supply and noises from household appliances such as the clothes washer and fridge. This receiver suffered from microphonics as well. The AF amp was a LM386 set for gain = 50.

A simple LC resonator tuned the mixer RF input and proved essential. I tried different ways to couple a 25 cm long wire antenna such as tapping the coil and an L- network, however, my main antenna peeve was severe hand capacitance effects. Although this receiver does not require a preamp for RF gain, the isolation seems needed. If you put your hand near the telescopic whip antenna of my Realistic Pro 2020 scanner, no noise or RF disturbance arises.

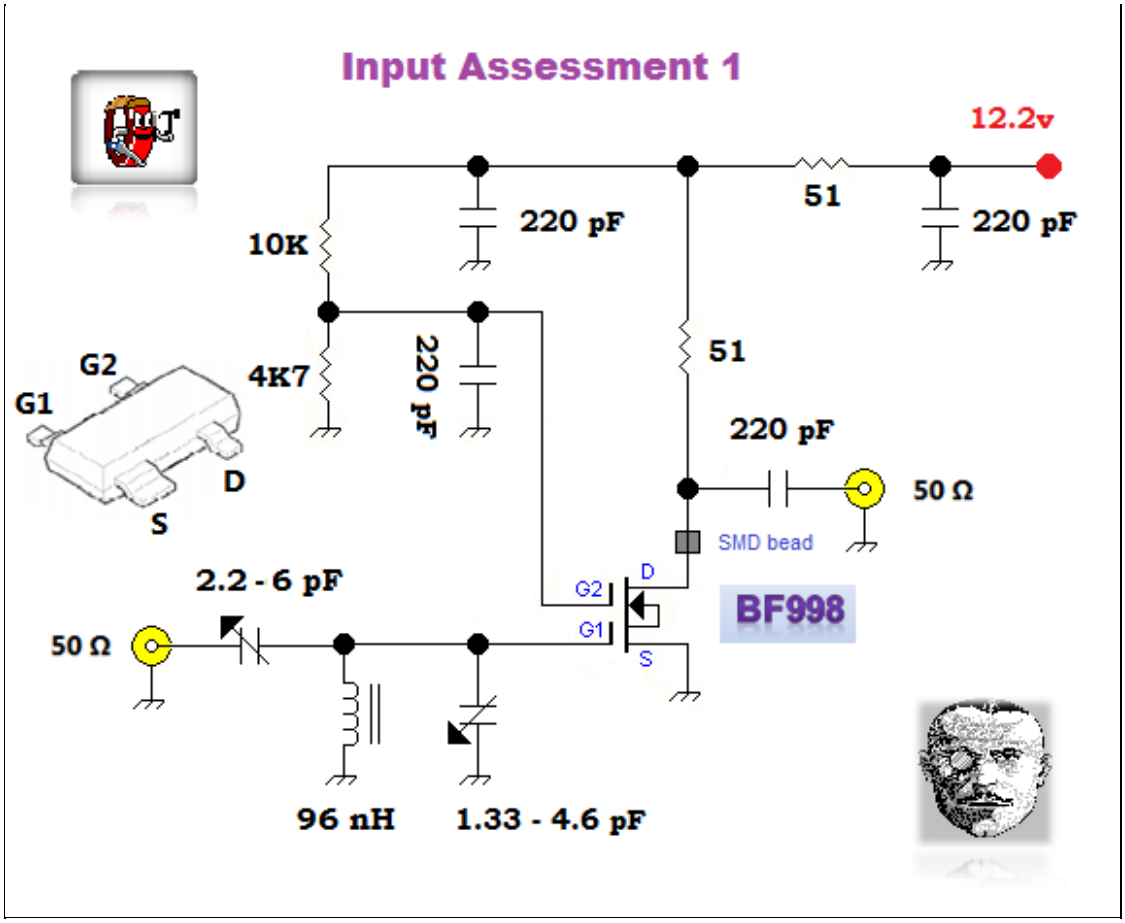
From studying scanner and FM receiver schematics, most rigs contain a dual-gate MOSFET preamp for a low noise figure and high input isolation. These MOSFET amplifiers feature input and output tuning, although with slug-tuned inductor cans and the oh-so-typical varactors, the resonator Q runs low.



Above — I quickly built a tuned-output common base amp to study front end isolation. I removed the existing resonator and connected this to the JFET mixer gate. It eliminated antenna hand capacitance effects. Click for a sound byte of the receiver with this preamplifier and a 20 cm wire antenna attached to its input. [206 KB download]. I reckon some of my receiver problems arose from the mess of unshielded boards, modules and wire on my bench.

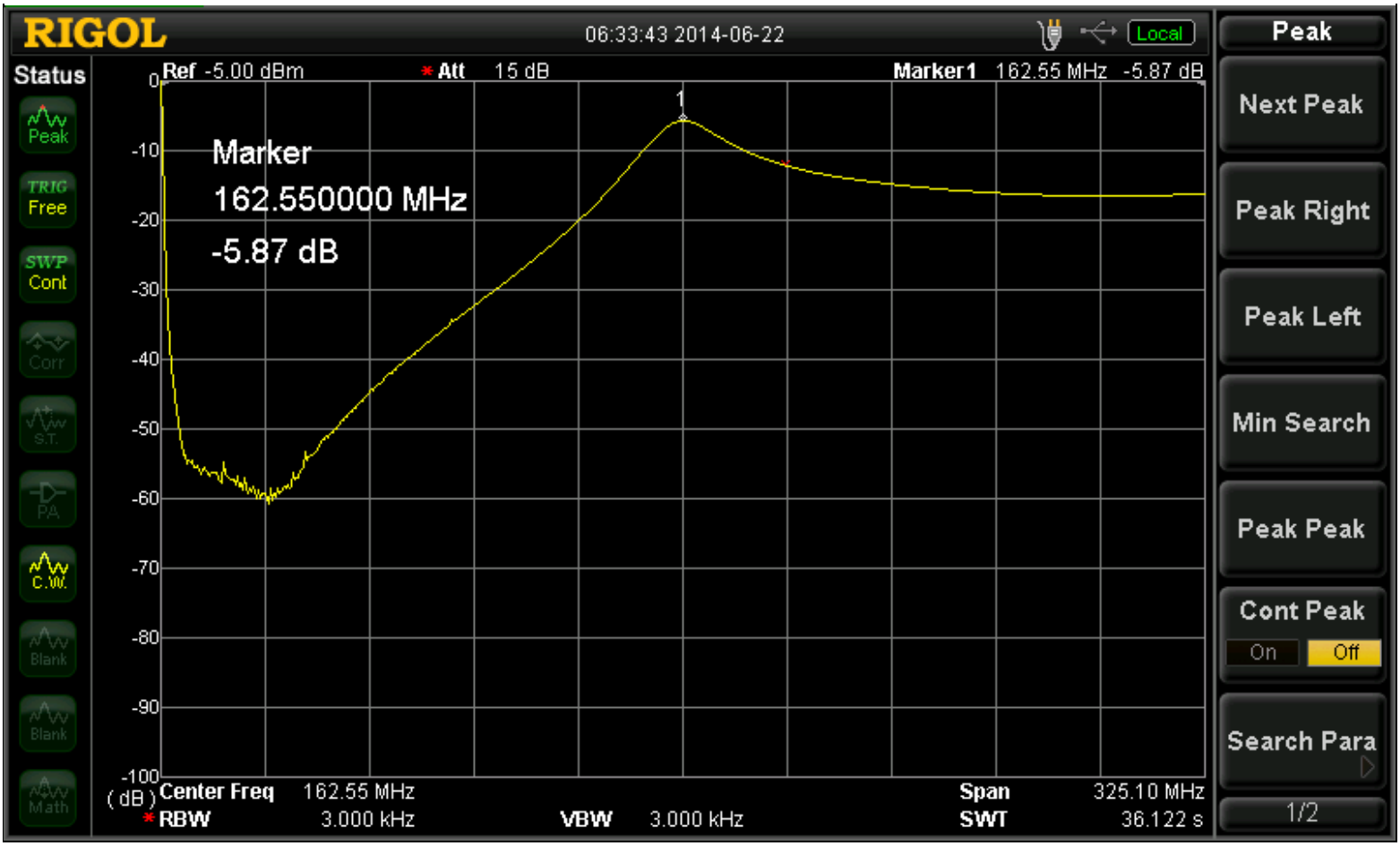
Tuned Input with a 2-Gate MOSFET [162.55 MHz]

A dual gate MOSFET preamp concluded this day's experiments. Tests of a simple input network follow:

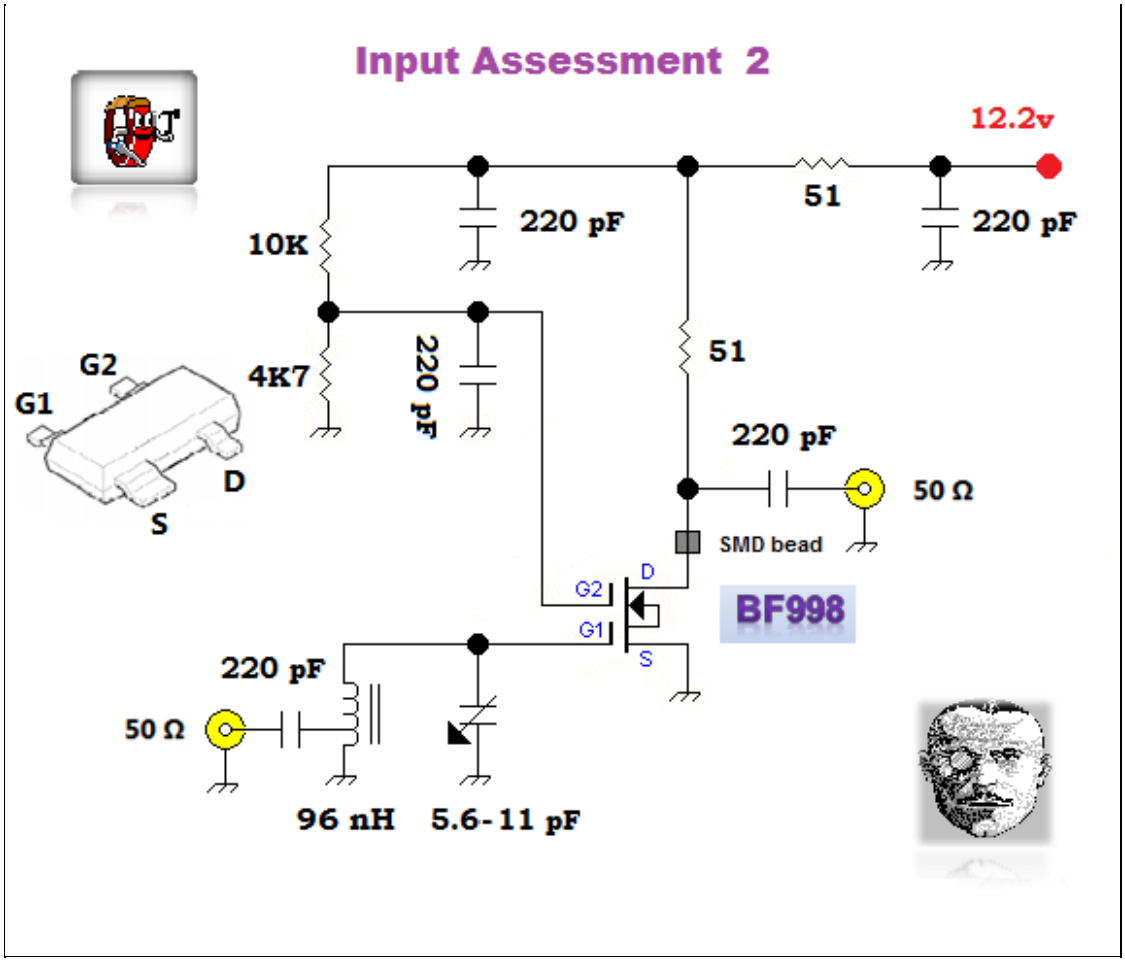


Above — Input Assessment 1. To examine the input without any output network effects, I loaded the BF998 drain with a 51 Ω resistor and matched the input parallel resonator for the greatest input return loss [-S21] by tweaking the series SMD trimmer capacitor. After setting the best input return loss [21 dB], I swept this amp.

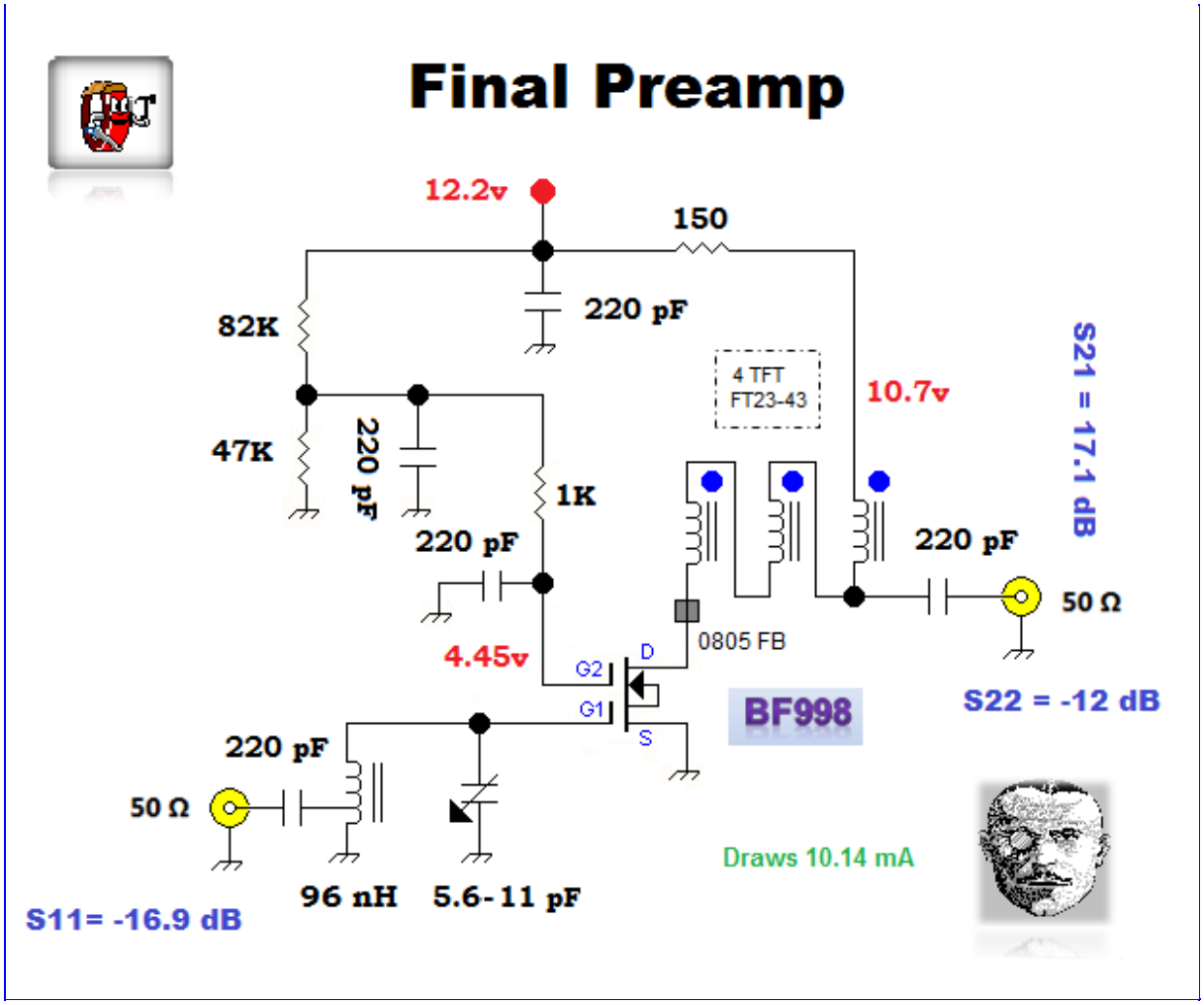
The XL of the 96 nH air wound coil = 98 Ω and the XL and XC of this tank may be scaled to other frequencies by keeping the reactance values shown or choosing your own. For example; a coil for 50 MHz where XL = 98 Ω = ~312 nH. I discussed and swept the ferrite bead soldered on the FET drain in Section 1 of the VHF Veronica web page.



Above — A sweep of the Input Assessment 1 amplifer. Although easy peasy to adjust, series matching the resonator to its input load results in a slipshod right-hand [low-pass] skirt. Click for a screen shot showing the amp's 3 dB bandwidth.



Above — Input Assessment 2. After measuring a poor low-pass skirt in Input Assessment # 1, I opted for the conventional tapped transformer to steepen the low-pass skirt response versus frequency. My sweeps *did* show a better low-pass skirt shape . Finally, I built a "real" amp and soldered in a broadband output network, tweaked the bias scheme + limited the VDD with a series 150 Ω resistor. A higher value input resonator trimmer cap is required for this amplifier versus that in Input Assessment # 1.



Above — The final 162.55 MHz preamp schematic + some measures. Click for a breadboard photo. Click for the gain measurement with a TG & SA. I measured the same S21 [within 0.3 dB] using my VHF signal generator plus my 50 Ω input DSO. I didn't push too hard with the gate bias or VDD, since neither maximal gain nor parasitic oscillations were sought. Click for a zoom of the filter peak response. Click for a wider sweep to see the right-hand skirt response.

Built on 2-sided FR4 with short cut paths to fit mostly size 0805 SMD parts, I placed via wires on every ground point and detected 0 unwanted oscillations from HF to 3 GHz.

A bare copper wire coil allows us to set the input impedance tap point anywhere along the wire — unlike the *integer only* tap points provided by magnetic toroid-based transformers. In LNA design, many builders set the input tap point for the lowest NF with the amp in noise figure test mode. Another vanilla flavored option is to set the tap with the amplifier input connected to a return loss bridge while adjusting the tap point for the best input return loss — that's what I did. My S11 = - 16.9 dB with my final tap around 1.2 turns of the total 7 links on my 96 nH coil.

The output transformer presents a poor match to 50 Ω with a S22 of just -12 dB. I plan to put a 4 dB pad after it for my NOAA receiver, so it's okay. Some might chase a better output return loss and also perhaps tune the output for better band-pass filtration.

Your FM receiver preamp might need better input filtration, however, if a simple design works, then simple rules! Have fun with *your own* experiments.

473
270
472

RF — Test and Measurement

Pin Outs

Mini-Circuits Labs SBL-1 and TUF-1 diode ring mixers

...more later

Above — Normal set-up: **Ground** -- Pins 2, 5, 6 and 7. **LO** -- Pin 8, **IF** -- Pins 3 and 4, **RF** -- Pin 1

Above — **RF** -- Pin 1, I**F** -- Pin 2, **Ground** -- Pin 3, **LO** -- Pin 4.

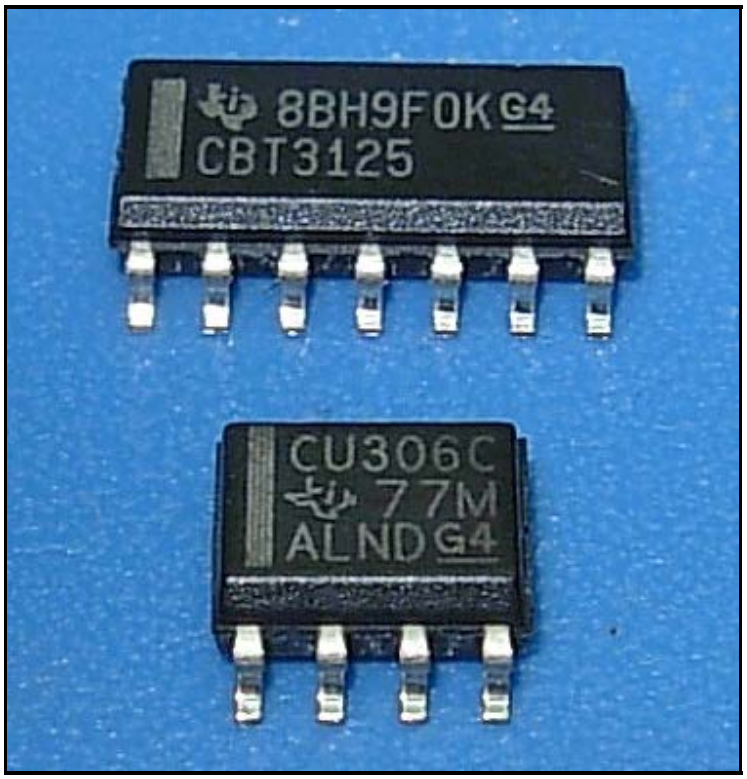
VHF-FM Supplement — Amplifiers

Deleted October 26, 2013

Amateur Radio Electronic Design

KL7R Memorial Receiver Experiments

Introduction



This web page is a memorial to Mike Caughran, KL7R, who died suddenly in January 2007. Mike was a passionate experimenter who was embraced by the homebrew radio electronics community. He was best known as the co-creator of and sidekick to Bill Meara, M0HBR on the podcast Solder Smoke. Mike was a hardcore science and technology buff. His knowledge of general science and curiosity about minimalist RF designs was amazing. I worked Mike on 40 and 80 meter CW and later by voice on eQSO. Mike was the first person to refer to this web site as the "popcorn site". He held an interest in digital circuits. In tribute to Mike, a series of receiver experiments which includes some digital circuits are presented. Mike Caughran will be remembered as a remarkable, kind and passionate homebuilder.

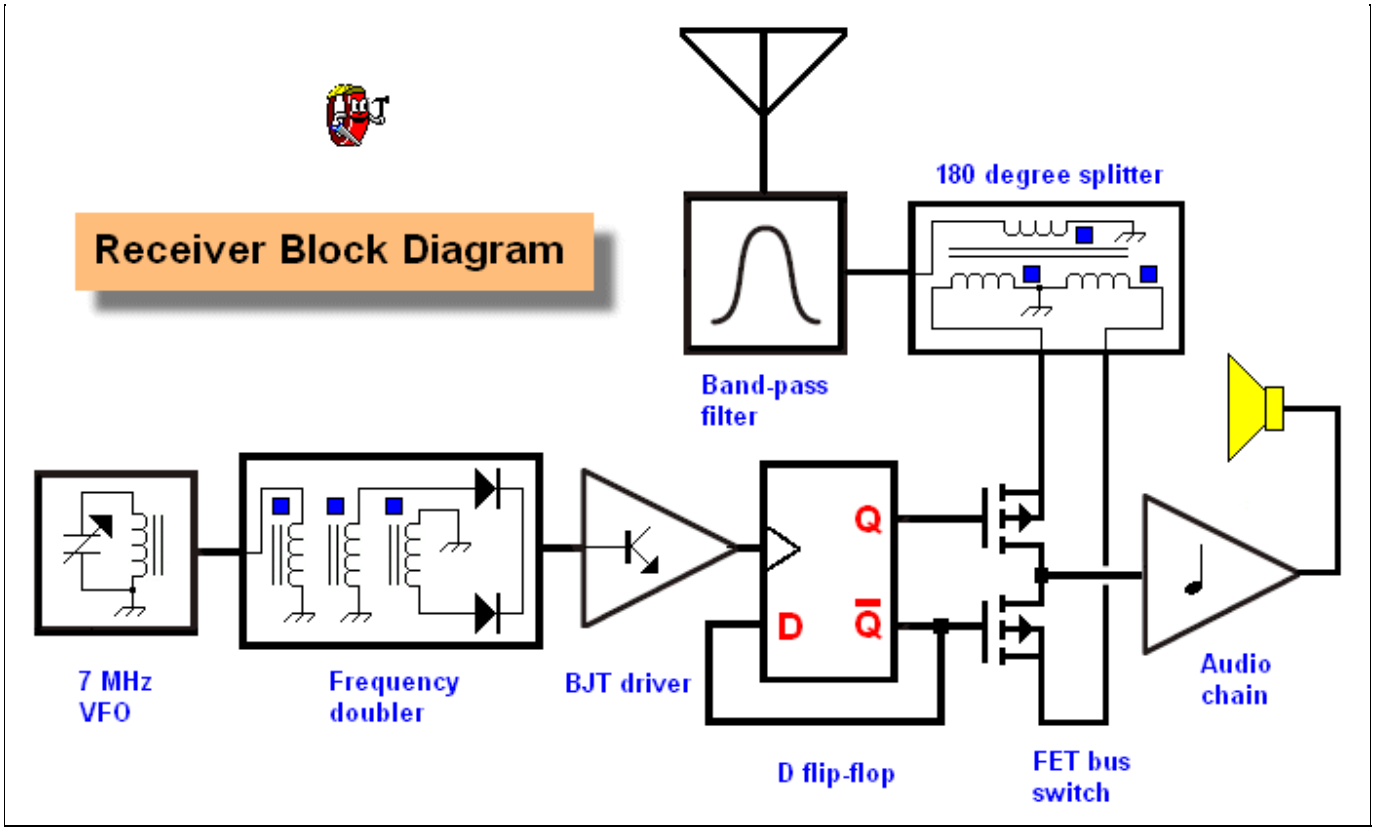
Navigation

There are 3 linked web pages associated with this KL7R tribute web page.

1. Mike's personal web page His own web page memorializes him best
2. Supplemental web page Additional schematics and photos which supplement this main web page
3. VFO web page Describes the VFOs used for these experiments

My special thanks to Wes, W7ZOI for his coaching and suggestions to improve many of the circuits on this web page. This web page borrows heavily from his designs as presented in EMRFD.

Receiver Description



Shown above is the receiver block diagram. The KL7R memorial receiver depicted on this web page is the final output of many hours of experimentation. Most of the circuits or circuit ideas originated in EMRFD.
From our conversations, Mike was always challenging himself; experimenting, testing and pushing his knowledge threshold. The joy of discovery motivated him. Fired by his spirit of inquiry, I explored methods to build a receiver containing at least 1 digital circuit. On many days, I accomplished nothing. The circuits did not work and little to no progress was made. These were the difficult dry spells all experimenters must endure. Design and circuit failures can be very disheartening. I also wasted a lot of parts. However, I kept going and slowly successes occurred and my confidence rose. The end result was a little more knowledge and a cool, popcorn, direct conversion receiver which I hope will provide ideas and inspiration for your own experiments.

Double Tuned Band-pass Filter

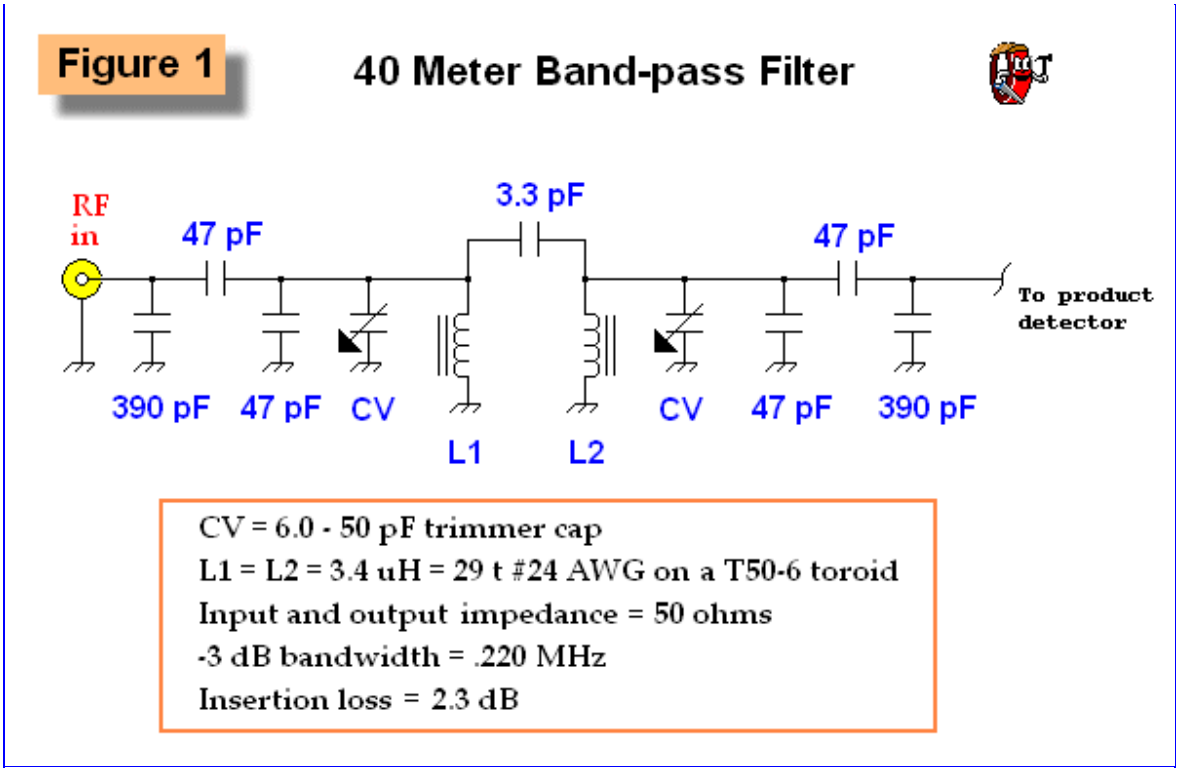
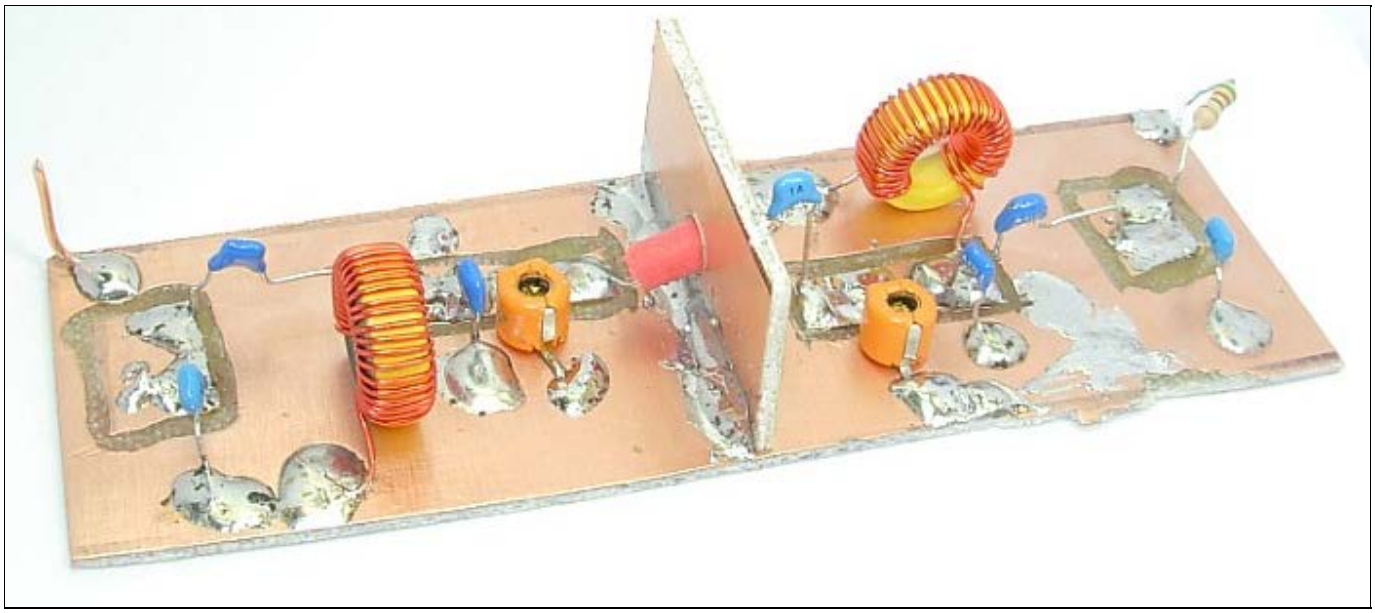
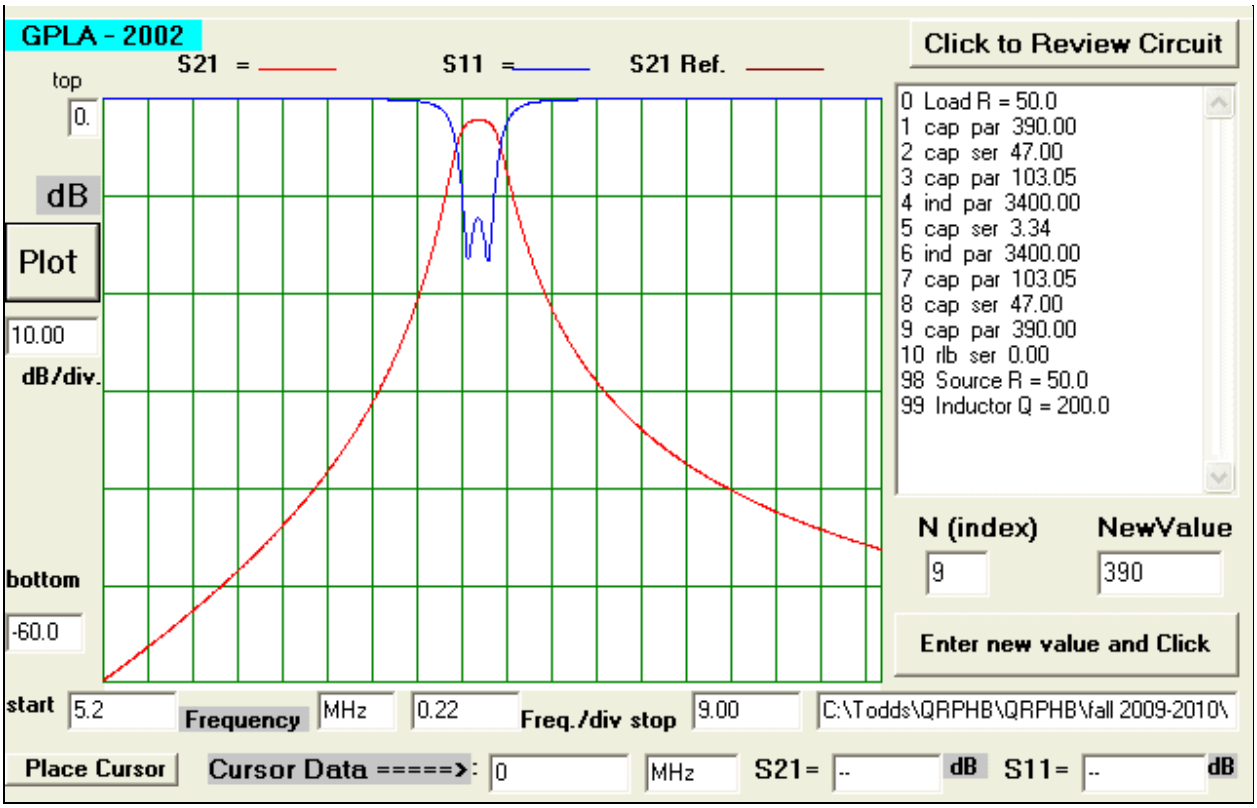


Figure 1 shows the front end band-pass filter. If you can't obtain a 3.3 pF coupling capacitor, try this other 7 MHz band-pass filter circuit or perhaps just design your own. See the Webmaster's page for information concerning many of the parts used on this web page.

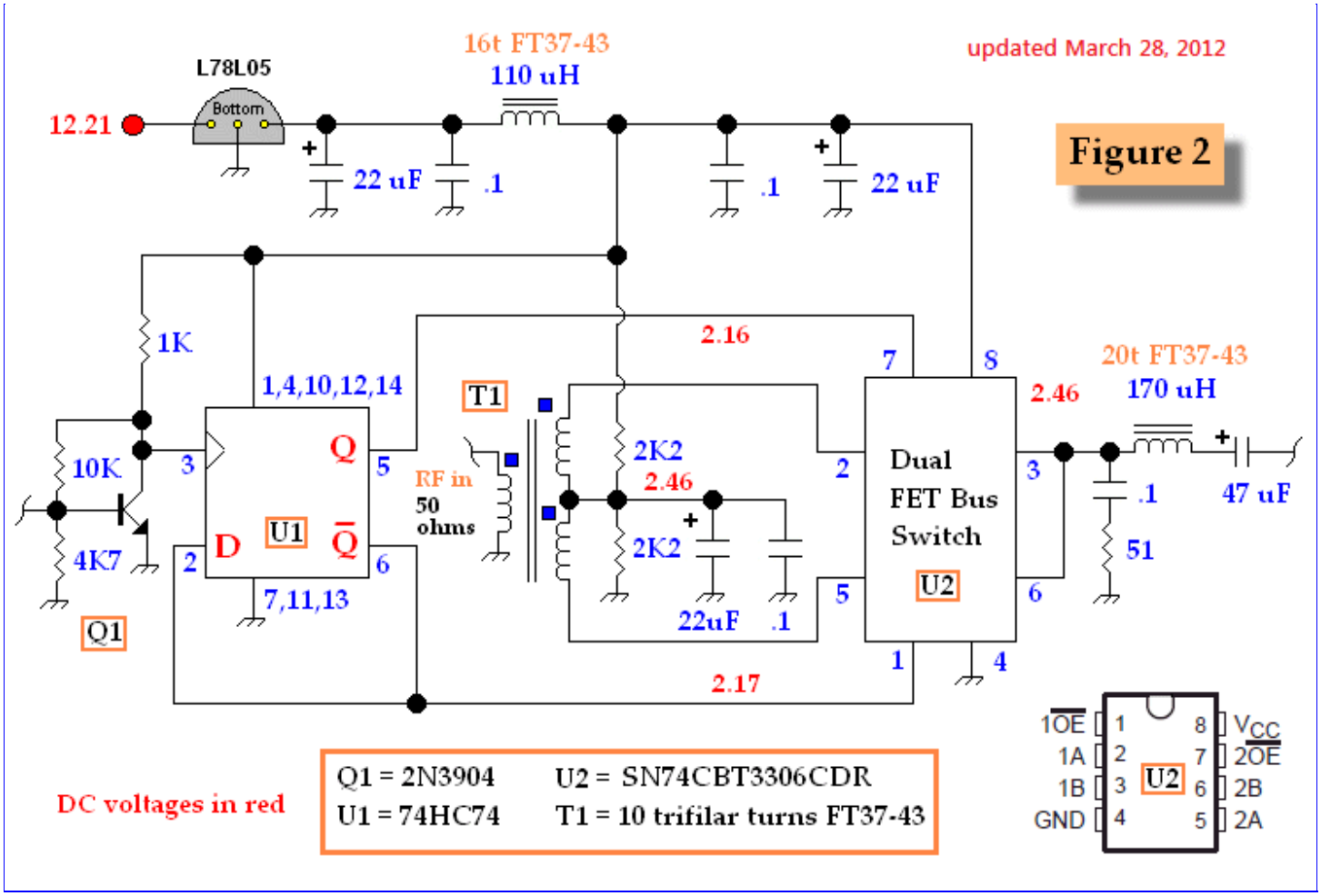


Shown above is a photograph of the Figure 1 breadboard. The inductors were spaced apart at right angles to reduce unwanted coupling. The copper clad board L-C tank divider is not necessary. The 51 ohm load resistor seen to the right was removed after testing.



Shown above is a GLPA simulation of the Figure 1 band-pass filter. Хорошо.

Product Detector Experiments

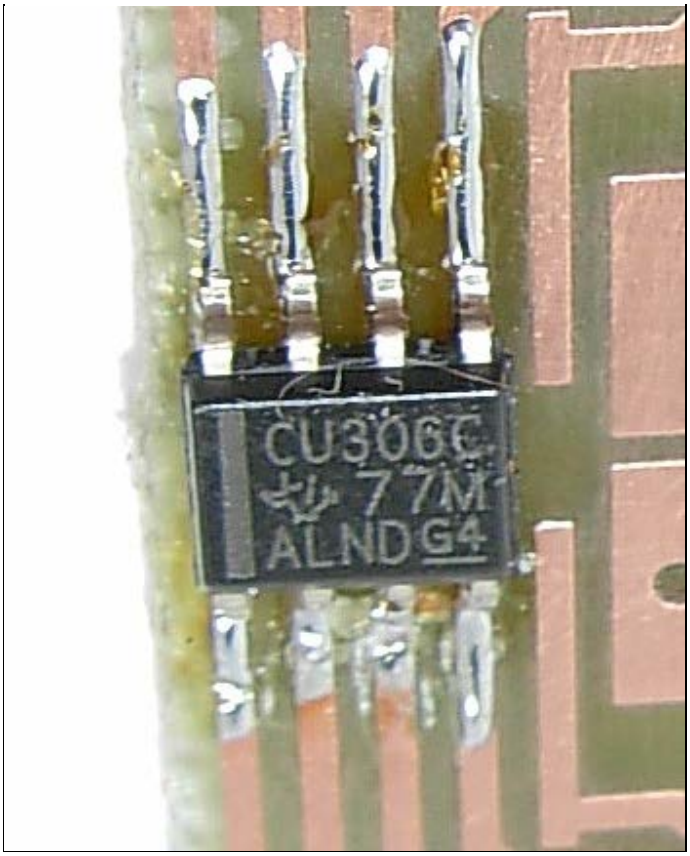


Shown above in Figure 2 is the BJT driver, D flip-flop and the CMOS switch product detector. The 14 MHz VFO connects to Q1 via a 0.1 uF coupling capacitor that is shown on the VFO schematic. A dual FET bus switch (CBT3306) serves as the product detector. The on-resistance of this switch is only around 3 ohms! If you are ambitious, you might try using a 14 pin SOIC switch such as the QUAD FET SN74CBT3125DR with 2 pairs of the 4 switches wired in parallel. I tried 3 different CMOS switches in the U2 slot. The other switches were the MAX4066CPD and a 74HC4053 (wired up appropriately using their datasheets). The insertion loss and performance of these 2 switches was disappointing. My bench standard for comparison was a 7 dBm diode ring mixer. Numerous experiments were performed. For example, I tried running the 4066 at 12 volts VCC to minimize its on-resistance and had to modify most of Figure 2 as well. Being new to digital, blending 5 volt and 12 volt logic IC required great effort to get it working properly. These experiments consumed the better part of 2 days. My conclusion was that if you are going to go to the trouble of make a CMOS switch work, you might as well use a part that has a low on-resistance. Hence, I have since abandoned using DIP IC CMOS switches (4053, 4052, 4066 etc.) as mixers and product detectors. They may still be a good choice in your own context.

The CBT3306 is outstanding and very similar to the diode ring mixer with respect to insertion loss and audio quality as a product detector. There are some other good CMOS switches you might try. I chose the CBT3306 because it was SOIC (the largest of the common SMT topologies), costs only 77 cents (Canadian dollars) and only has 8 pins to deal with. Pragmatism on the workbench is always good!

SMT versus VE7BPO

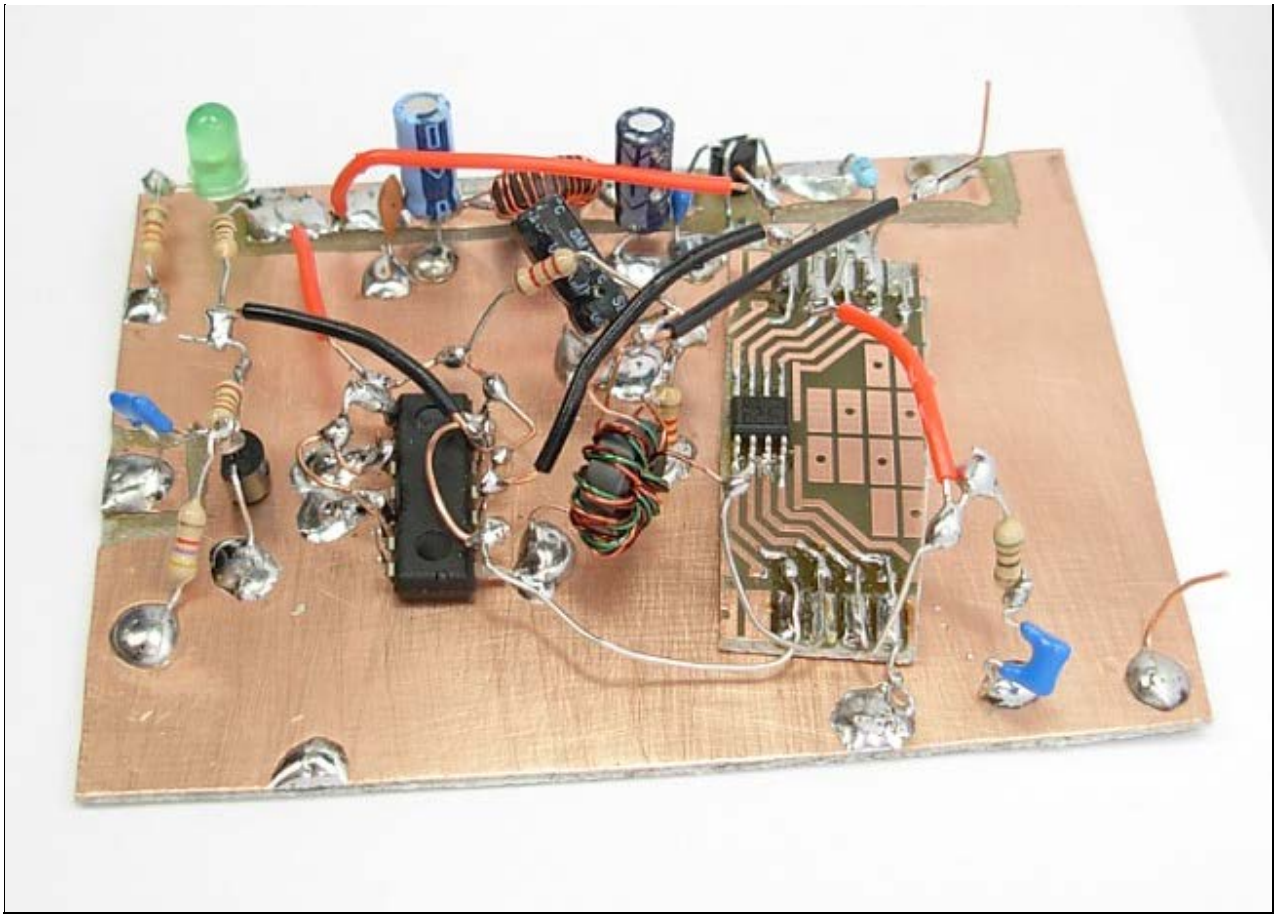
The difficulty with using the CBT3306 was that I had to learn about and equip for SMT. I ordered the switches and after their arrival, hesitated to do anything with them for 5 days. I managed to borrow a magnifier and bought a SOIC prototype board and some flux. Still, I seemed to be paralyzed with fear about soldering U2. I was stuck. For inspiration, I went back to Mike's web site and found this web page. Here Mike was working with a 18 pin SOIC chip and I was worried about a mere 8 pin IC! Part of the problem was my relatively poor eyesight. I learned this can be managed with a magnifying visor. I bought mine here , but only after U2 was soldered in. The borrowed table magnifier was okay, but constrained arm movement and reduced lighting. The visor seems to be a better choice.



Shown above is the CBT3306 soldered on my prototype board. I put flux on the board traces and then tinned them. Following that, I lined up U2s leads on the traces and began soldering. The bottom pins (pins 1-4) did not go well, but they were soldered all the same. Following that, the soldering of pins 5-8 went very well. I was very happy; the SMT monster had been tamed! It was learned from Wes, W7ZOI that many builders use Surfboards for SOIC applications. I will get some for future SMT IC work. Using SMT parts in your Ugly Construction allows you to use parts which are unavailable otherwise. Increasingly, good old DIP ICs are disappearing from catalogs and some new parts are appearing as SMT only. It makes sense to jump in and use SMT parts when it is advantageous for spec reasons, or if miniaturization is required.

With some effort, the Figure 2 circuit could be morphed into Colin, G3SBI's H-mode mixer format. Not on this web site though! If you Google "H-mode mixer" , many good websites will be returned.

The product detector's baseband audio output is at the 47 uF capacitor which connects to Q1 of Figure 3. A simple 51 ohm, low pass network is used to terminate U2. This is from W7EL's An Optimized QRP Transceiver from QST for Aug 1980. I have some more information regarding AF termination circuits on this web page. You may wish to increase the 170 uH inductor value somewhat to get more low pass filtering. The product detector circuit is the number one potential source of hum and noise in this receiver. Take the time to plan your layout to minimize wire length and crossing and provide some physical symmetry. Decouple well.



Shown above is the product detector. At this point, I had not decide how to terminate U2 and had soldered a shunt 0.1 uF cap and 51 ohm resistor to the switch output. U2 is dwarfed by the FT37-43 toroid; a part we normally think of as small in size.

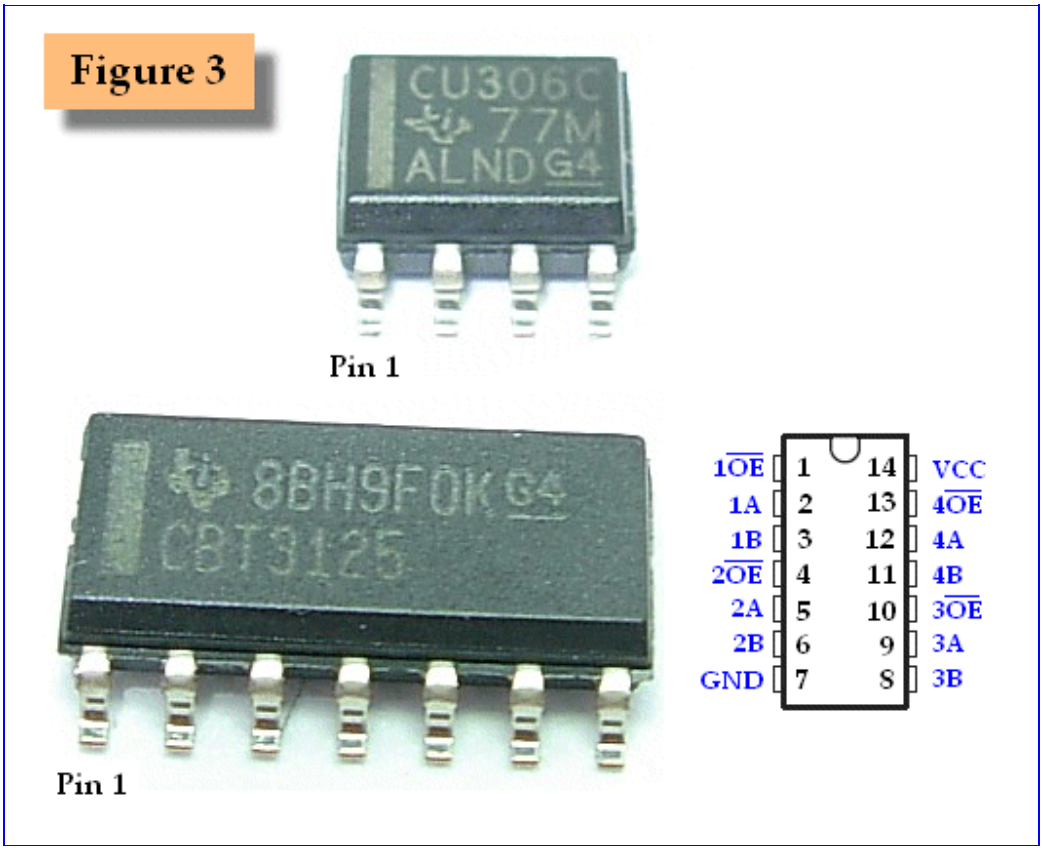
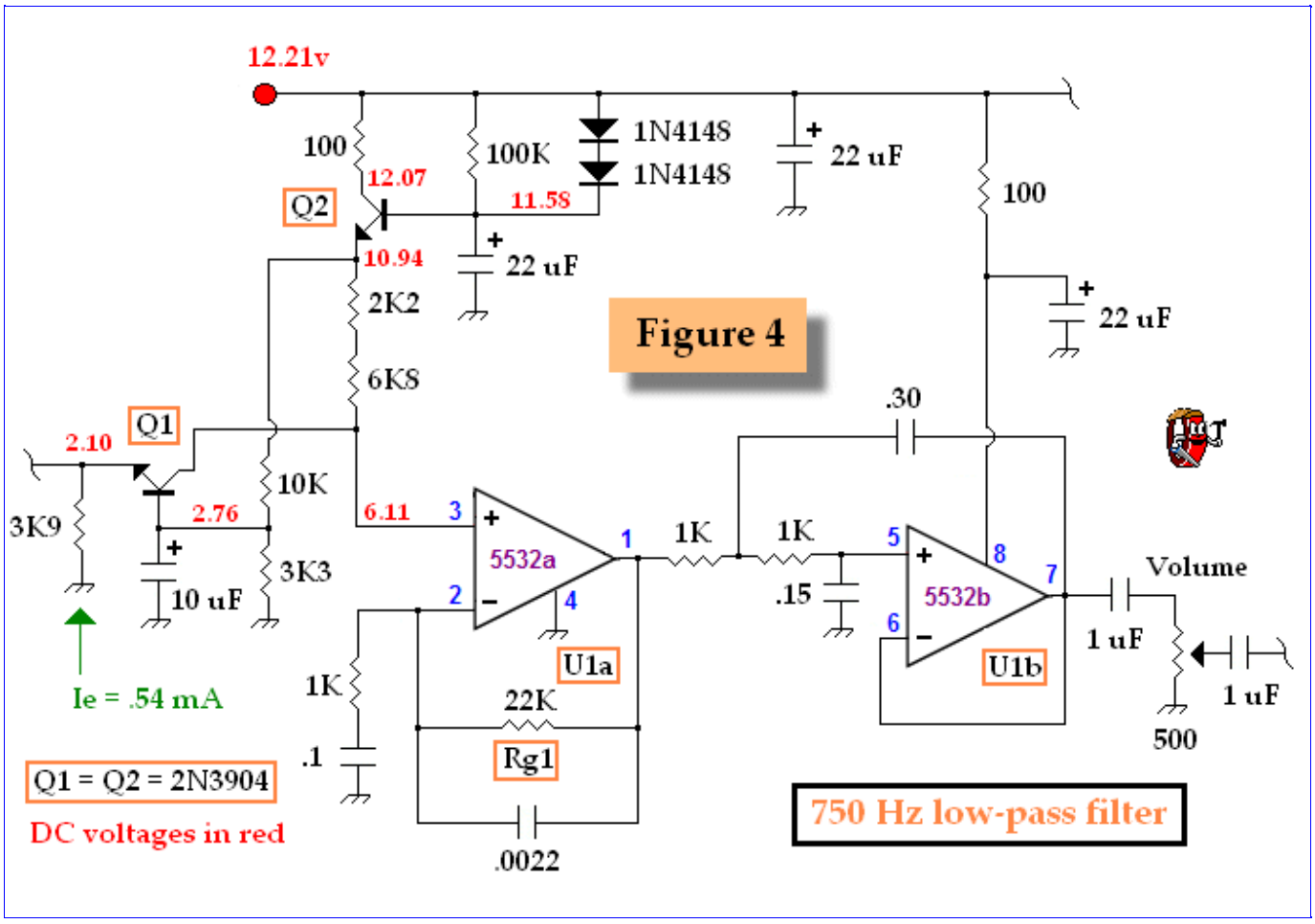


Figure 3. Photographs of the FET bus switches I have in my collection and the pin-out for the 14 pin CBT3125

A diode ring mixer was used as a reference mixer in the product detector experiments.

Audio Amplifier Chain



Shown in Figure 4 is the audio preamp and first low pass filter. Direct conversion receivers are all about audio. High performance DC receivers have become more common since Rick, KK7B unleashed his R1 on the world in 1992. High performance receivers are out of scope for this popcorn website, however, this design is welcome. The entire audio chain uses low-cost 5532 op amps, cheap BJTs, plus fairly common resistor and capacitor values. Poly"something" capacitors were used for all audio AC coupling and shunting capacitors of 4.7 uF or less.

For hum immunity, the familiar BJT capacitive multiplier popularized by Roy, W7EL was used at Q2. Using a high-end spectrum analyzer, Wes, W7ZOI demonstrated that this circuit can oscillate. He detected oscillations at UHF. Thus, Wes recommended using a 100 ohm snubber resistor on the Q2 collector to alleviate these parasitic oscillations. Additionally, 2 diodes were added in parallel to the 100K base resistor. This provides an instant-on feature for the audio preamp, as normally there is a time-constant delay when you switch on the receiver. I like this feature, but it is purely optional.

Q1 is a common base amp biased for 0.54 mA. Therefore, the input impedance is 26/0.54 or about 48 ohms. The 6K8 + 2K2 collector resistors were paralleled so that the quiescent collector voltage was close to 6.1 volts. You could also substitute a single 9K1 resistor, although this is an uncommon value. This voltage provides the approximate VCC/2 bias needed for U1a and U1b. Connecting the Q1 collector directly to the pin 3 op-amp input allows the exclusion of a coupling capacitor and the usual VCC/2 resistor network used to bias the 5532 op-amp from a single power supply. I borrowed this from EMRFD.

The gain of U1 is set by the resistor labeled Rg1. If after testing, the AF gain is too high (for example if this AF stage was used in a superheterodyne receiver), simply lower this resistor value. I chose a 22K ohm resistor to allow enough gain for weak signal listening. On louder stations, you will need to lower the volume control to prevent distortion in the stages that follow as overall, there is a lot of gain in this AF chain. Adjust the Rg1 value to suit your needs. If you use a switch such as a 4052 for U2, or your receive antenna is small, you may want to increase this resistor a little for more gain. This is an experimenter's receiver after all. The 0.0022 uF capacitor in the op-amp feedback loop provides a single pole of low pass filtering. I ran this capacitor as high as 0.0082 uF. This gave a theoretical 3 dB cutoff frequency of ~ 880 Hz, but the receiver lost its sizzle. Experimentally, I learned that using a 0.0022 uF feedback capacitor on both U1A and U2A dropped some of the high frequency noise while preserving some sparkle in the received audio.

The audio preamp chain is concluded with a 750 Hz low-pass filter. Resistor values were kept low to minimize noise. All the active low pass filters are low Q, Sallen-Key designs with a Butterworth response. They are stable, easily scalable and brain-dead simple. These filters are fatigueless due to their gentle cut off slope. I actually had a 0.30 uF capacitor in my parts collection, although it was a 600 volt rated part from my tube amp parts bag. You could use two 0.15 uF capacitor in parallel or just substitute a 0.33 uF capacitor. If you can't find a 500 ohm volume control potentiometer, swap in whatever you have, but try to keep the resistance low.

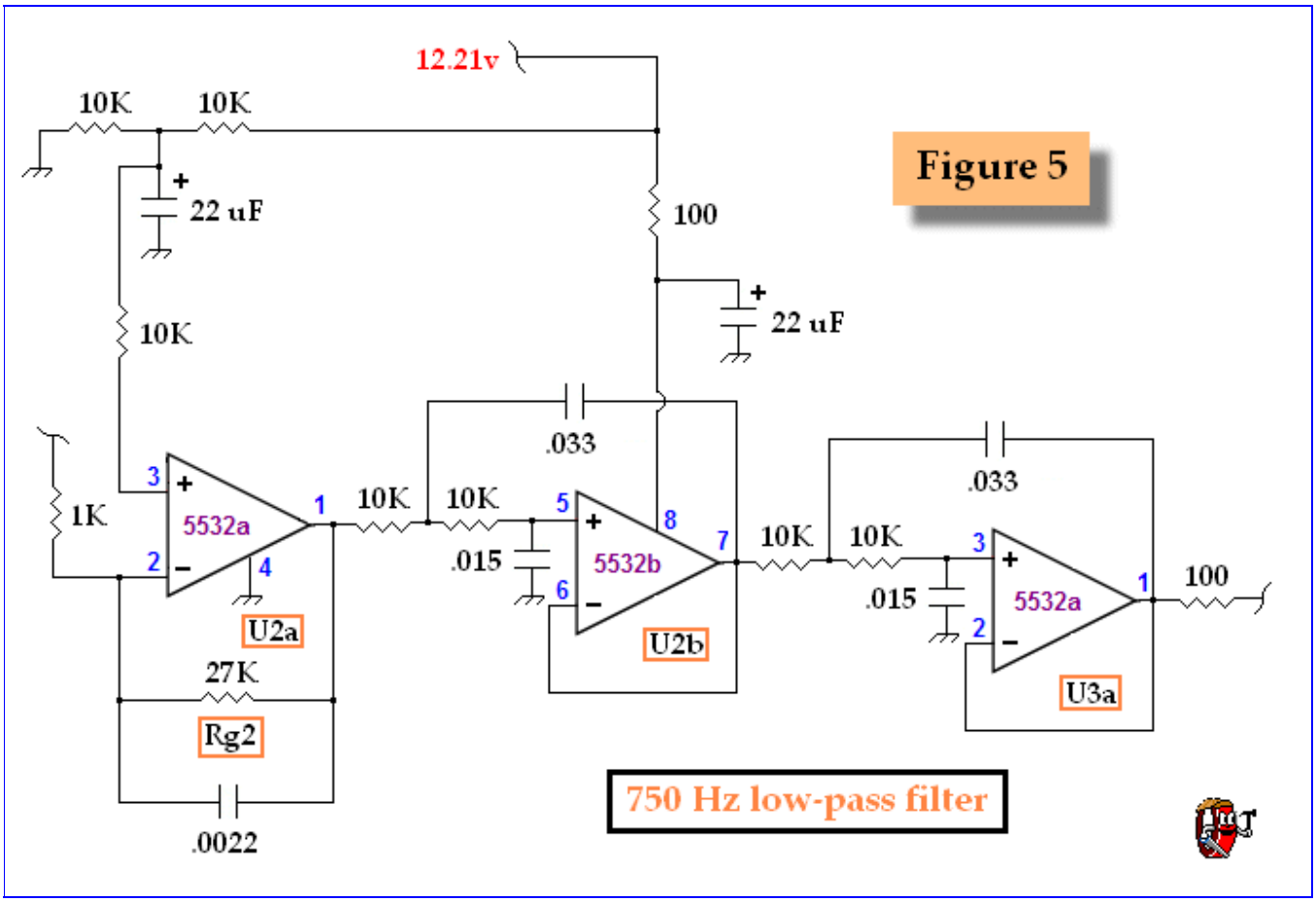


Figure 5

In Figure 5, the remaining preamplifier stage and low-pass filters are shown. Rg2 was chosen for the same reason as Rg1 in Figure 4. Please experiment with these values. The low pass filter stages are scaled up by a factor of 10 as resistance values are less critical at this point in the audio chain. You could use the same resistor and capacitor values used in the first low-pass filter or use the Figure 5 values for all of the low-pass filters. Very often, our parts collections dictate how we experimenters design and build circuits.

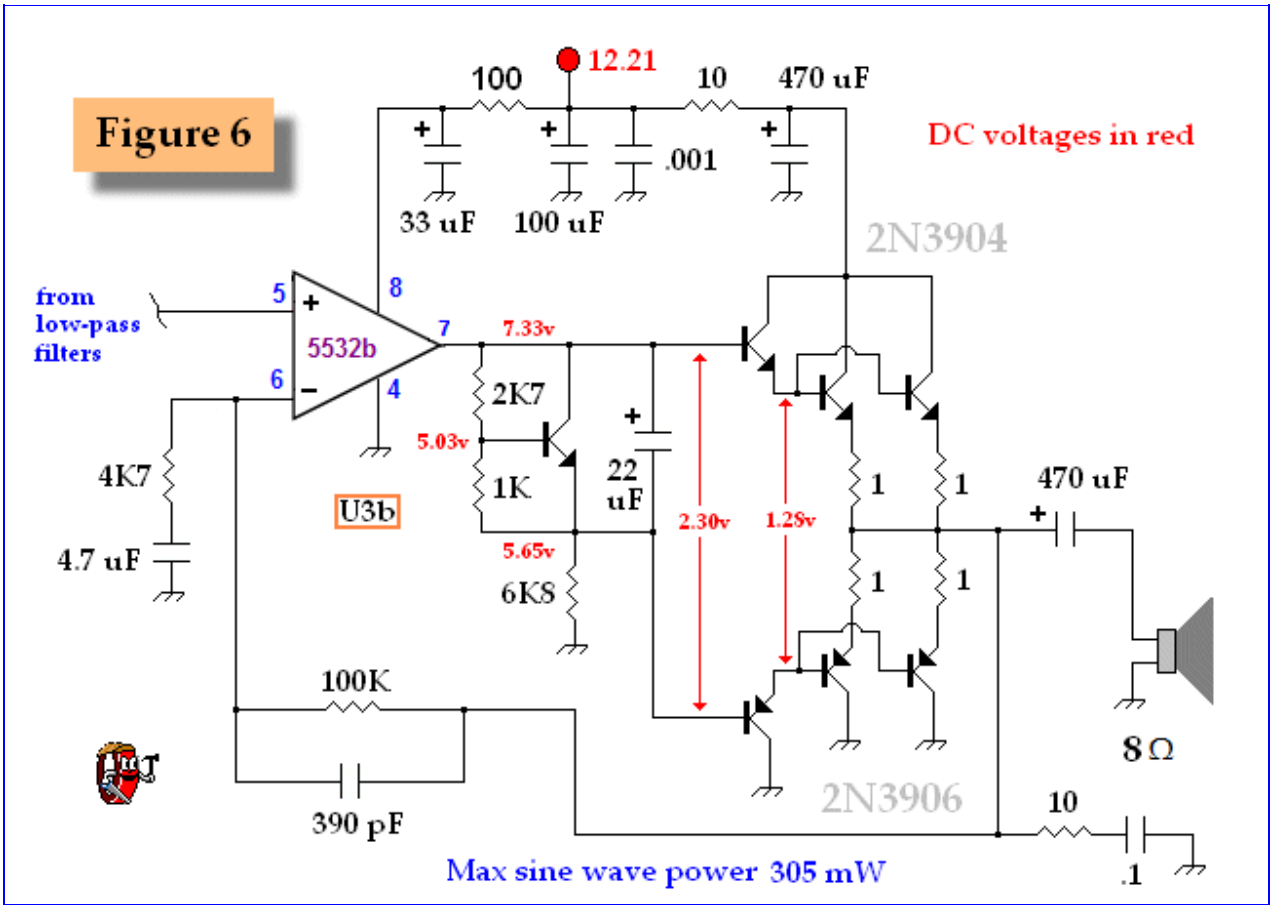
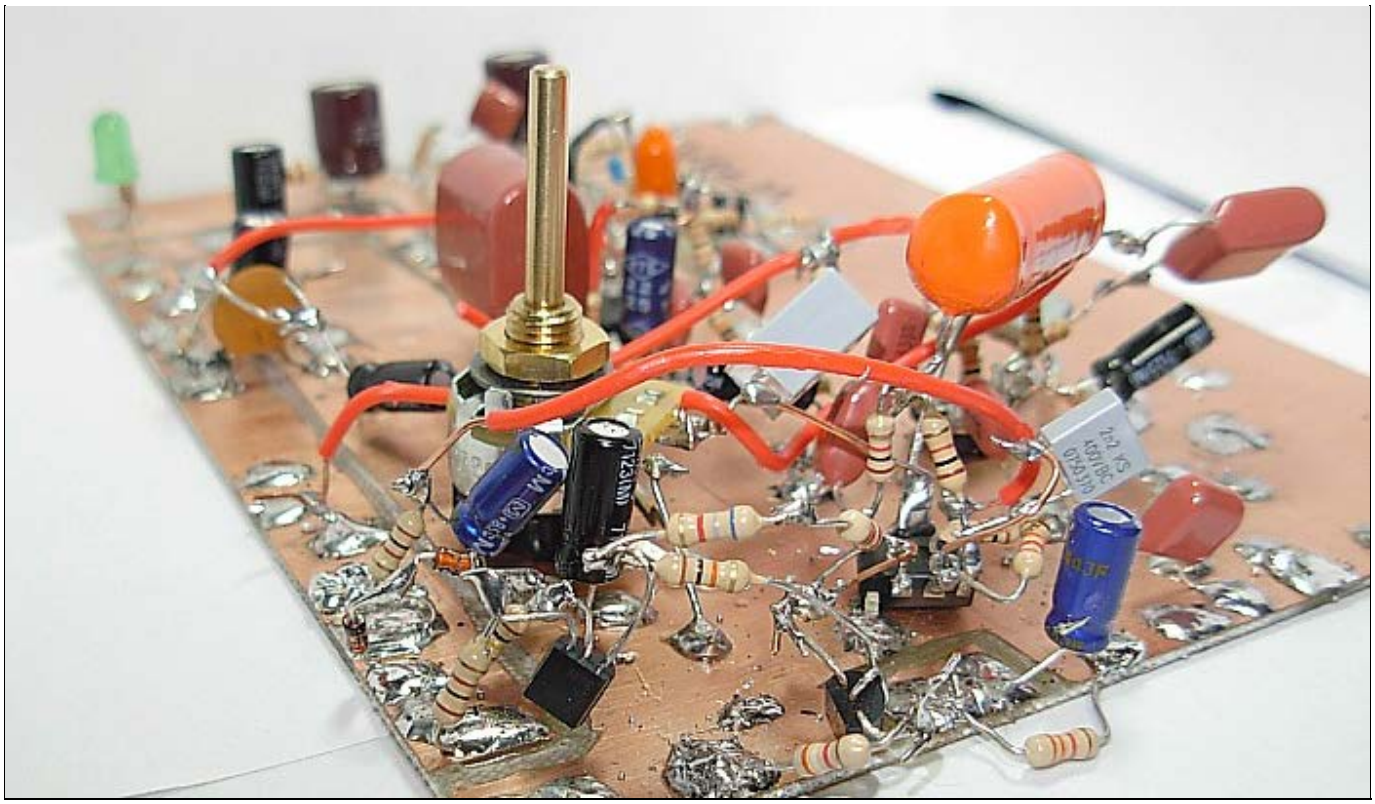
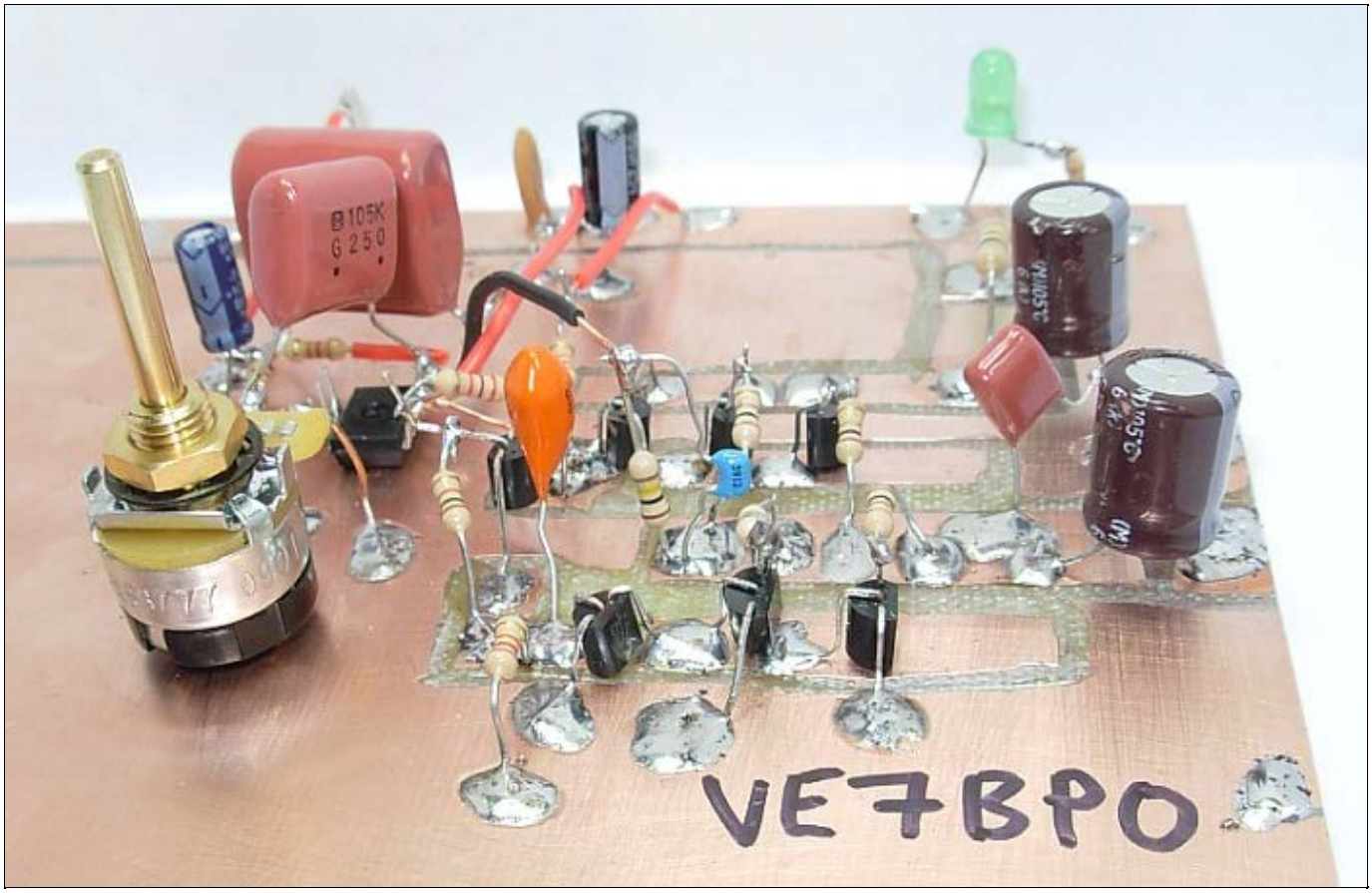


Figure 6 is the power amplifier schematic. Apart from field portable transceivers where headphones are used to save battery power, I exclusively listen to my receivers via a loudspeaker. This safely allows the exclusion of AGC circuits. You may have noticed that AGC circuits have not been presented on this web site. I rarely use them. The extensive VCC decoupling in Figure 6 is needed in this high gain AF chain. The 1 ohm emitter resistors in the finals invite instability, but were tamed with the 10 ohm / 0.1 uF low pass filter. My original design called for 2.2 ohm emitter resistors, but none were available. This final amplifier is moderately loud and very quiet. When it was first built (the AF chain was built backwards), you could not tell it was on when no signal was applied. With no antenna connected to this receiver and the volume on full, there is only a little noise. This is a good way to test a receiver AF chain for noise. Speaker choice is also important. Speakers of a greater power rating and size sound better; especially when mounted in a wooden cabinet.

Biasing power amps has been discussed extensively on this web site. Ensure you measure and record your quiescent DC voltages as shown in Figure 6. If you hear cross over distortion or the quiescent voltage between the bases of the paralleled final transistors is less than 1.25 volts, try decreasing the 6K8 resistor to 5K6 ohms.



The final preamp breadboard is shown above. You can tell by all the grunge and the solder marks, that many different configurations were trialed. Despite looking haggard, the AF chain is quiet and does not hum nor detect broadcast radio.

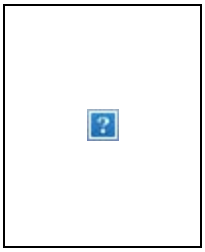


Shown above is an early breadboard of Figure 6. In this version, the volume control was at the input and a voltage divider network was wired to pin 5 to provide VCC/2 bias. Testing with an audio signal generator, a tape player and other sources were performed. Later, the biasing resistors and the potentiometer were removed and the Figure 5 stage was added to the copper clad board and tested.

Final Thoughts

We are remembered best by those whom we affect. Certainly Mike's unique perspective and enthusiasm inspired me to dig deeper into this hobby. I asked Bill, M0HBR, for a quote to conclude this web page. Bill wrote this: "Just last week somebody was asking me for background info on Mike's DSB modification of the Heath HW-8... I think it is a real tribute to Mike that years after his passing, hams around the world are still talking about him fondly, still visiting his web site, still following his lead on homebrew radio projects. Certainly among the 3,000 + listeners of the SolderSmoke podcast, Mike is present in spirit every time a soldering iron is heated up." Thank you Bill.

My heartfelt regards to Mike's family.



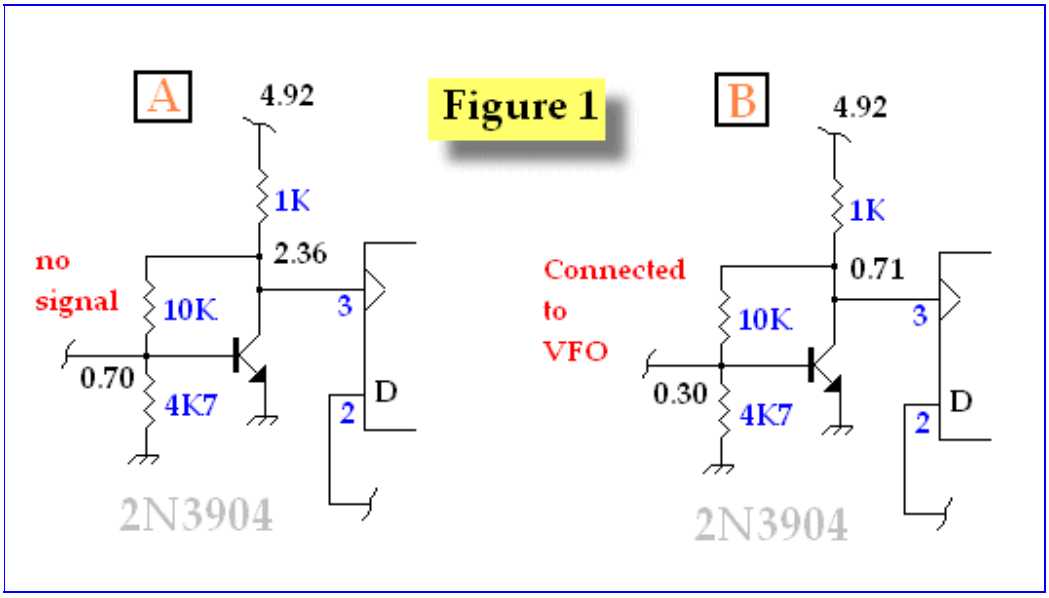
VFO Experiments 2009

Amateur and Short Wave Radio Electronics Experimenter's Web Site

Supplement to the VFO 2009 Web Page

This web page is a supplement to the VFO 2009 Web Page

Figure 10 Bipolar Transistor Notes



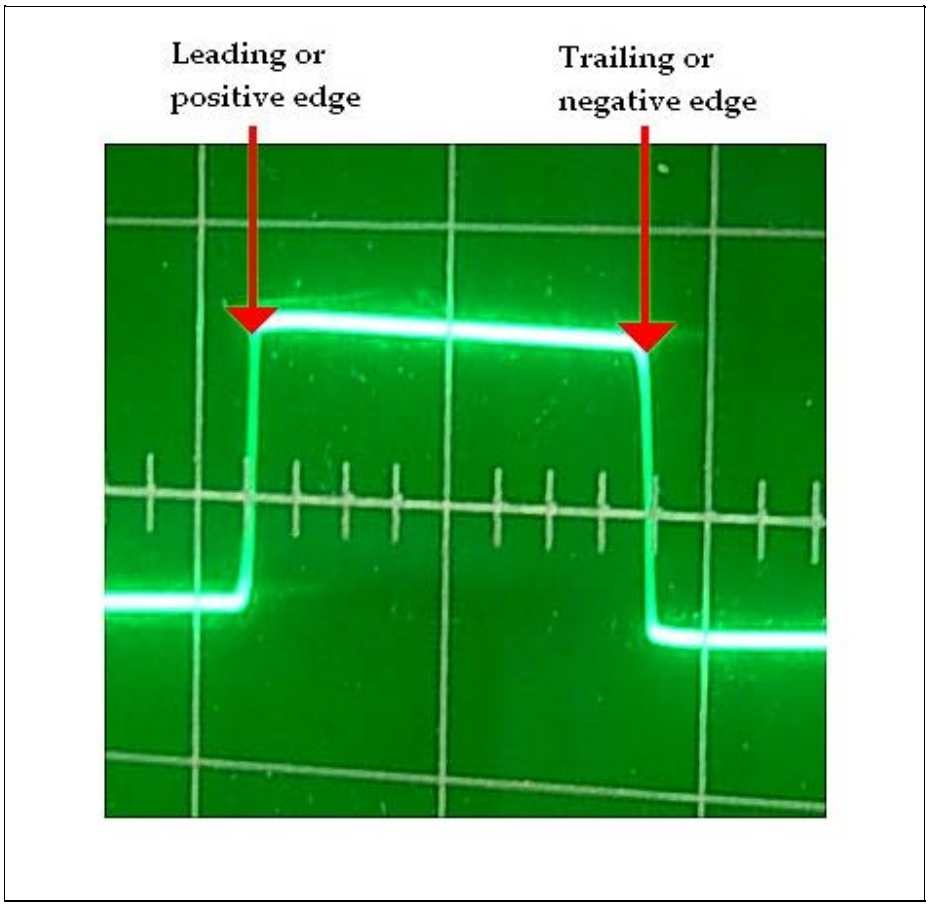
This circuit is from EMRFD and is an analog to digital interface for the VFO. I did not understand the circuit, so I asked Wes, W7ZOI to explain his design criteria and the basic circuit function. This transistor circuit is a driver. The purpose of this transistor driver is to positive edge trigger a D flip-flop. The paramount criterion was that the transistor collector has a quiescent bias voltage that fell between the minimal acceptable logic high and logic low DC voltages on the 74HC74 clock input. See this chart for details about this topic. The correct DC collector voltage is set via feedback from a voltage divider (in this case, a 10K and 4K7 resistor). The collector DC voltage (shown above in Figure 1A) was 2.36 and lies perfectly between the logic low and high state of the 74HC device that it drives.

This transistor circuit is also a high gain amplifier, however, its primary function is a driver. Refer to figure 1B. When connected to the VFO output, the base voltage drops to 0.30 volts. This is an average DC voltage because it fluctuates as the AC waveform swings up and down. More or less AC drive on the input will change this DC voltage.
When connected to the VFO output, the collector voltage dropped to 0.71 VDC. Again, this is just an average, as when the AC swings positive, the the collector voltage will drop down to saturation (where collector voltage is less than base voltage) and then increase towards the positive supply. It does not have to go into cutoff; just to 3.2 volts or so. It also does not have to go into saturation; just to below the 1.3 volts or so. Thus, the transistor remains between cutoff and saturation when appropriately driven with a sine wave.

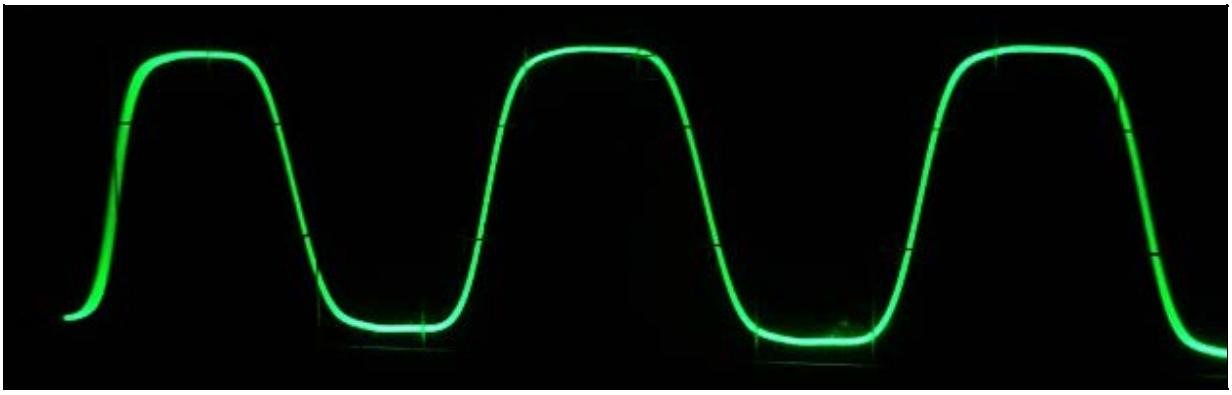
Our VFO serves as the clock and data on the D flip-flop inputs are only transferred to the flip-flop's outputs on the positive edge of a clock pulse. Positive edge triggering means that the output only responds to the input changes when the clock signal transitions from logic low to logic high. What a fantastic little circuit! You can find additional information and some practical examples in EMRFD. My special thanks to Wes, W7ZOI for

helping me to better understand his design.

Positive Edge Triggering



The triggering of a positive edge triggered flip flop is better understood viewing a graphic like above. The instant the waveform goes positive, the flip flop is triggered. It is easy to understand why a square wave is desirable for clean triggering.



Amateur and Short Wave Radio Electronics Experimenter's Web Site

Radio Amateurs of Canada Support Page

Commentary for Canadian Radio Amateurs

A significant number of licensed Canadian amateur radio operators do not support their national amateur radio society: Radio Amateurs of Canada Typically support is by membership, but may also include volunteering time to perform the various functions required to run a large, non-profit organization. It has been demonstrated repeatedly, that a small percentage of people are doing most of the work in modern volunteer organizations.

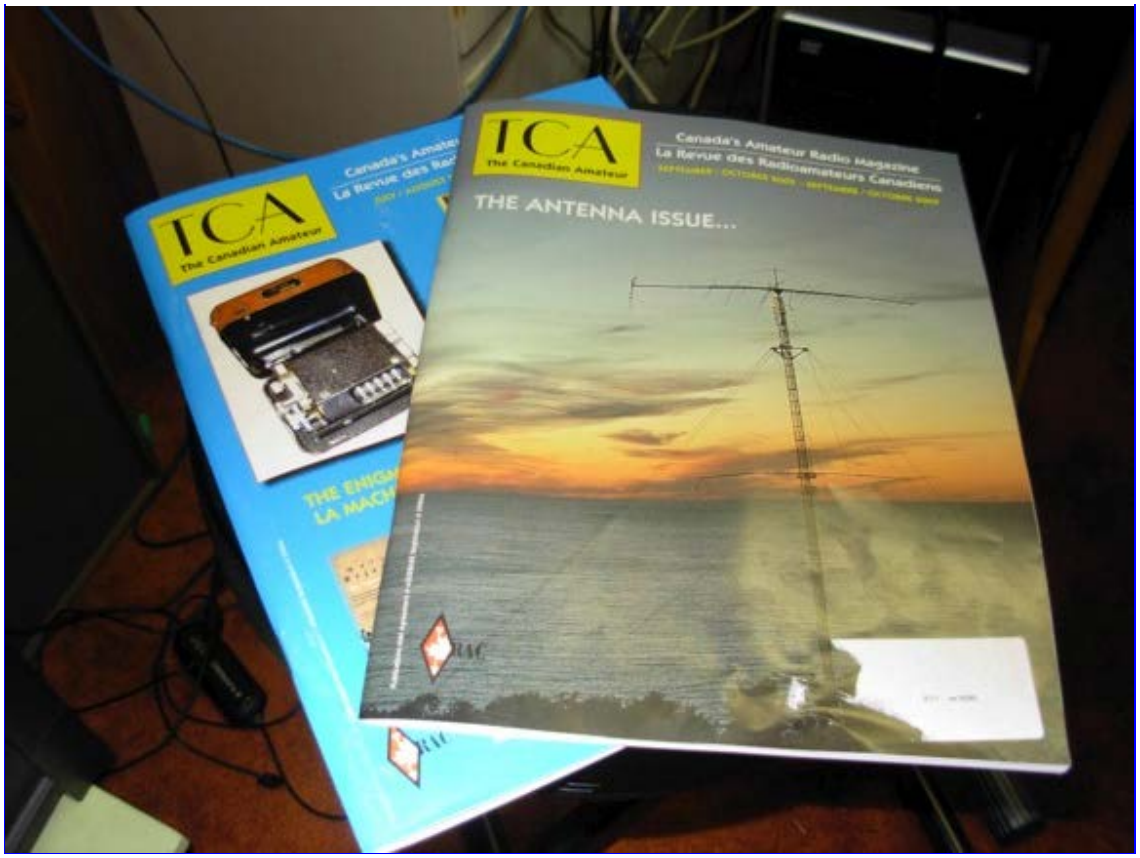
Some thoughts concerning volunteerism

- Volunteerism is down in many sectors including service clubs, meal delivery programs, church groups, volunteer firefighters radio clubs, etc. etc.
- The population is aging and our current volunteers are retiring; Younger folks have so many distractions and things they can do today. We now live in a world with 200 plus TV channels! The spirit of volunteerism seems to be decreasing in the western world.
- There is some evidence that people who donate their cash to charities also tend to participate in groups, associations and organizations. Is this a characteristic or trait?
- People volunteer for causes they believe in. What do people believe in now? Is our culture too self-absorbed, individualistic and focused on our own personal agendas and pleasure?
- Economics. Are young to middle aged people less able to volunteer and contribute because they are spending more of their time working?

RAC Membership

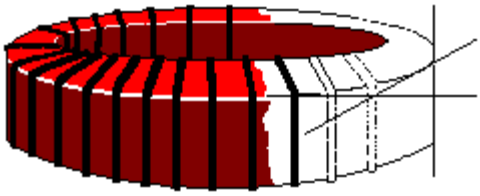
If you haven't, please consider joining the RAC. I joined out of respect to the many RAC volunteers who have devoted their personal time and continue to toil to keep amateur radio viable in Canada. We cannot take our currently allocated radio frequency bands for granted. There are groups who for profit, want to apply technologies which will consume or interfere with the radio frequency bands we enjoy. Canadian amateur radio needs a strong, united voice to survive into the future.

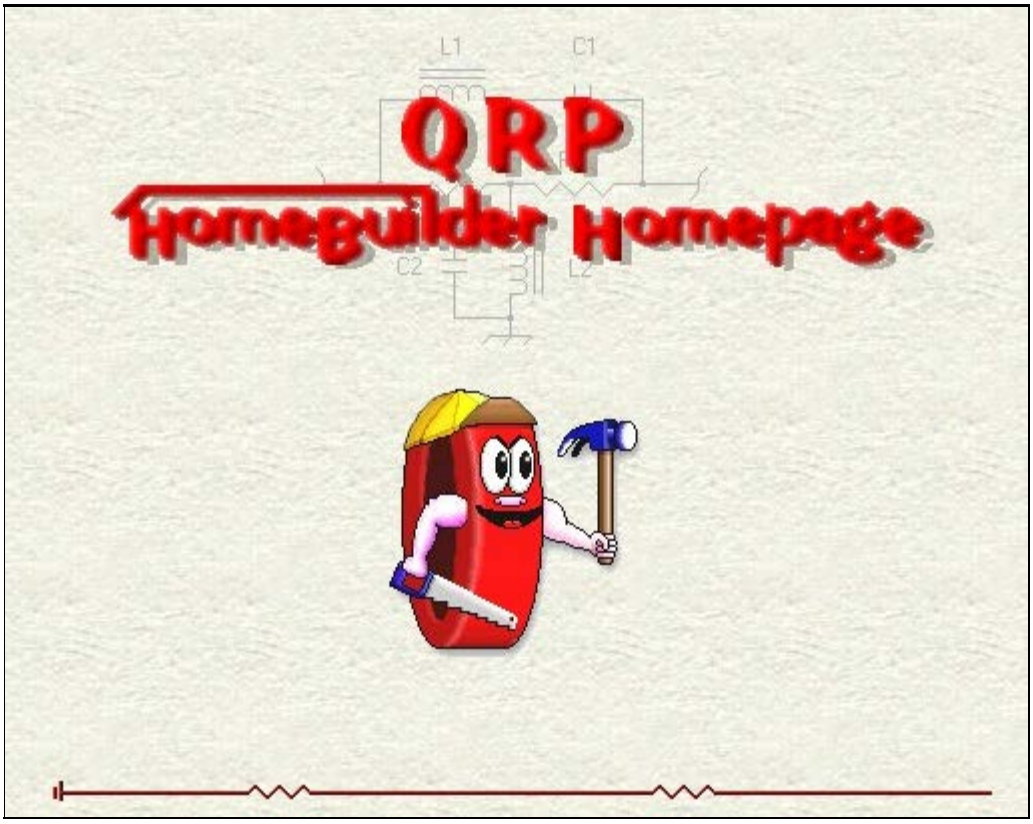
Apathy and lack of awareness may significantly decrease the rights and privileges we enjoy as radio amateurs today. The least we can do is support our national amateur radio society. It would be even better if more of us contributed by volunteering our personal time towards our wonderful radio hobby and serving the RAC.



The RAC Journal The Canadian Amateur is an excellent resource. Sept-Oct 2009 marked the first ever antenna issue.

Old QRPHB Web site graphics...

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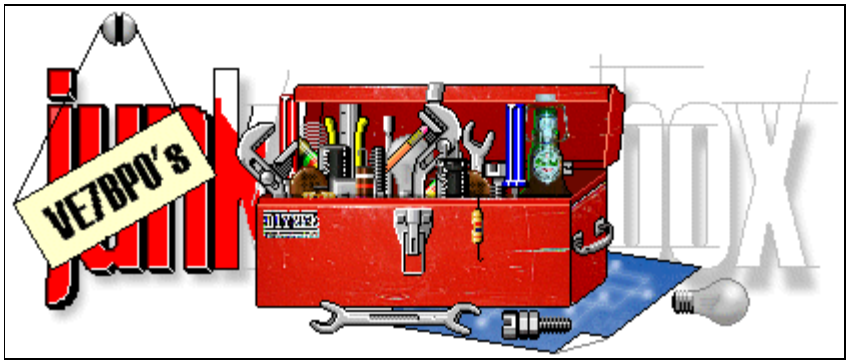
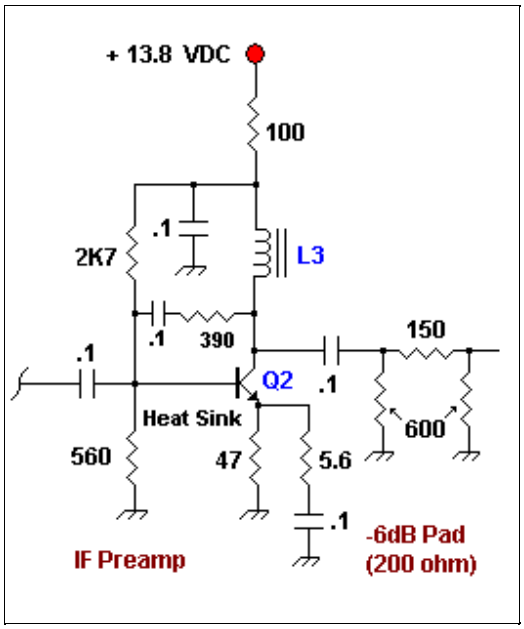

VE7BPU's

This page last updated: September 21, 2013

Info on the Popcorn Superhet RF Amp

Discussion



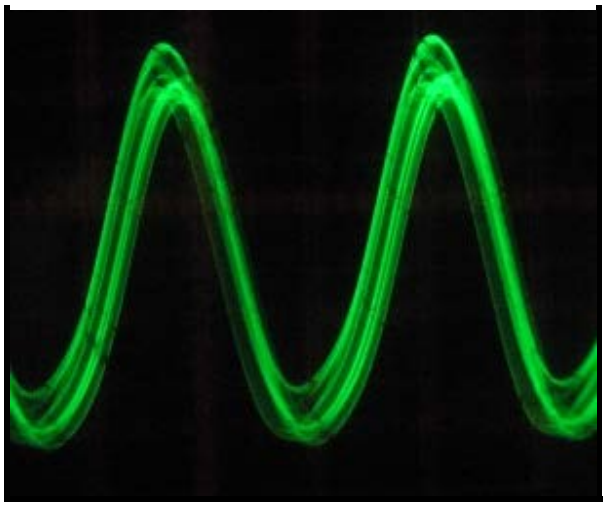
I have received a lot of emails regarding this RF amp from the popcorn superhet receiver. This amp was designed using a calculator and as it turns out, my input and output impedance was off by several ohms from the target 50 ohms input Z and 200 ohms output Z. Today we have computer programs to calculate the resistor values, so crunching the math is not a factor any more. This RF amp is preceded by a 50 ohm diplexer and followed by a 200 ohm pad, so it actually matches the preceding and following stages very well considering that it is clearly depicted to be a popcorn receiver. If I had to redesign this amp in 2008, I would use software to design a CE amp using "noiseless" Norton feedback, Still, this was a good sounding receiver at the time.
Probably the weakest section of this receiver is the audio preamp from Q5 on through to the LM386. I would likely never use these noisy stages in 2008-9. Instead I would use low noise op amps. This web site like my own abilities has grown over time. In the following section, I have paraphrased some writing from Wes, W7ZOI concerning the Q2 RF amp.

An Analysis by W7ZOI

The input impedance of a common emitter bipolar transistor amplifier is related to the standing current, the emitter degeneration resistance, the transistor beta, the F-t, and the operating frequency. Most of those details can be well modeled with the hybrid pi model. If you put quite a bit of emitter degeneration in such an amplifier, the gain drops while the input impedance increases. However, it is much less dependent upon beta and frequency. The emitter degeneration is one form of negative feedback. But now we introduce a second form of negative feedback by applying some of the collector voltage to a resistor that generates a current that is fed back to the base. This feedback tends to reduce gain just as does emitter degeneration. But it also tends to reduce input impedance. By playing a careful balancing act, you can design an amplifier that has a 50 Ohm input and output impedance with a reasonable amount of gain. This particular circuit has a gain of 21.7 dB at 10 MHz and an input return loss of 14 dB. The output return loss is even better at 18.6 dB. Z-in is 39-j14 while Z-out is 172-j34.

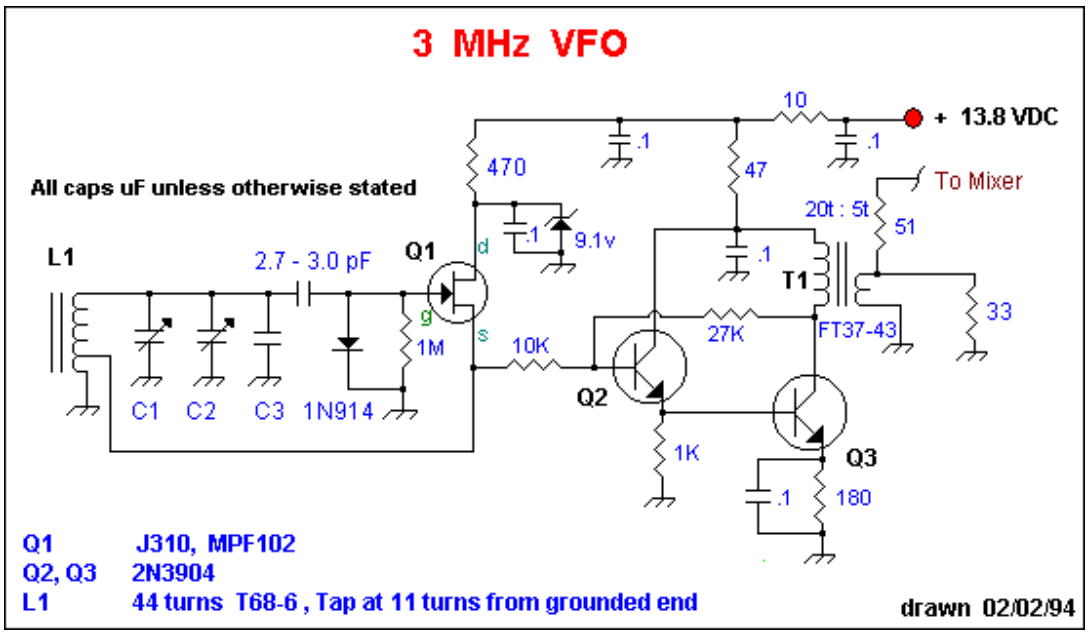
These are calculated values using that hybrid pi model. I have devoted a few pages to a discussion of this feedback amplifier topology in the ARRL 2003 book, Experimental Methods in RF Design. This was co-authored by KK7B and W7PUA. You can get the book on line from the ARRL. The book includes computer programs that will handle the design chores. One deals with bias issues while another does the RF design and analysis.

The RF amplifier under discussion may be found on this web page



3 MHz VFO for 40M Popcorn Superhet Receivers

Discussion



This VFO was first popularized by Roy Lewallen, W7EL and has been used in countless homebrew and commercially sold QRP rigs. For the 40 meter ugly super hets, you have a choice of a VFO frequency of 3 MHz or 11 MHz and I chose the former for this website. For L1, you can use the T68-6 as shown or redesign your own inductor using a T50-6 core. The T68-6 core allows the builder to wind the coil using # 24 AWG wire for enhanced Q and stability. The main tuning capacitor, C1 that was used had a built in reduction drive and went from 4 - 19 pF. If you use a tuning capacitor with a greater capacitance swing, you may have to connect it to the top of L1 via a small-value NP0 capacitor to reduce the tuning frequency range. This can be done using math or by just plain experimentation. C2 is a small ceramic air-variable trimmer from my junk box used to set the lower band edge of the VFO. It can be omitted if the user wants to go frugal and experimentally set the lowest frequency of the VFO using small-value ceramic NP0 caps. Air variable caps for C1 and C2 are mandatory for minimal drift VFO operation. C3 actually refers to 4 NP0 ceramic caps which were used to place the VFO on the correct frequency. Four caps were used to minimize heating and thus drift in the VFO. For the prototype VFO which tunes something ~ 3.00 to 3.67 MHz , I used the mentioned air-variable, four NP0 ceramic caps (20pF,100pF,5pF,100pF) and C2 the trimmer cap that went from 2 - 50 pF. These values should be used only as a guide for prospective builders as there are many variables at play. Following a 10 minute warm up period, the VF0 frequency stability is excellent. Again, keeps all component leads as short as possible. T1 is a broadband transformer wound by using 20 turns of # 26 AWG over a FT37-43 toroid core and then distributing 5 secondary turns spaced evenly over the primary windings. Do not omit the 33 ohm load resistor. The 2.7 to 3.0 pF coupling cap should also be of the NP0 ceramic type. Q1 can be the J310 (my favorite) , 2N4416 , MPF102 or other correct substitutes.

This VFO should be in a separate sealed, shielded box from the rest of the receiver

.

VFO Stability Hints

Numerous tips for enhancing VFO stability have been generated from Wes Hayward, W7ZOI, Roy Lewallen, W7EL, Doug DeMaw, W1FB and others. Here are a few summarized:

1. Use air wound or powdered-iron toroidal inductors made from number 6 material.
2. Use the heaviest gauge of wire possible to wind your inductor.
3. Anneal the inductor by boiling it in water for around 5 minutes after winding.
4. Use ugly construction above a SINGLE-SIDED copper side up ground plane.

5. Capacitors in the L-C circuit should be NP0 ceramic type.
6. Use air variable capacitors for tuning that have double bearings and no backlash.
7. The VFO should be operated at a lower regulated voltage.
8. Do not use cheap low-Q trimmer caps.
9. Completely encase the VFO to prevent RF leakage and to minimize environmental temperature changes.

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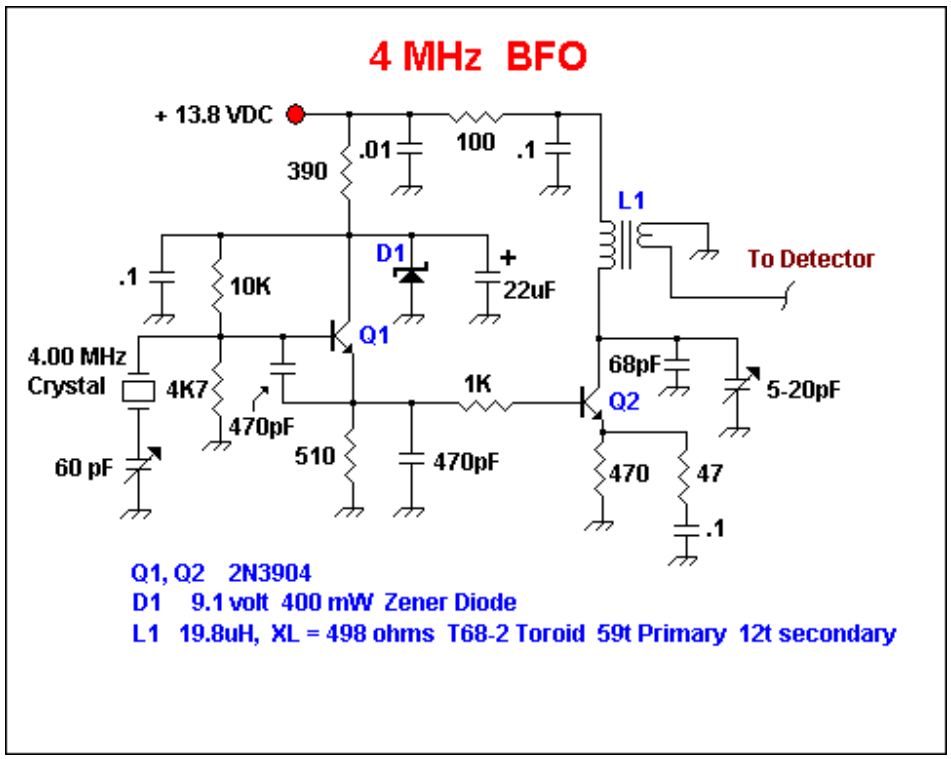
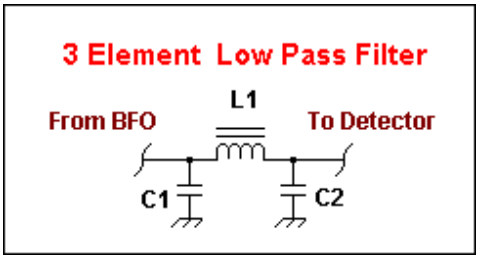
4 MHz BFO for 40M Popcorn Superhet Receivers

Discussion

Shown is a 2 stage BFO for use with the Ugly 40 meter superhets with a 4.00 MHz IF. This is a design by W7ZOI which I have used from 3.5 to 9 MHz by only changing the crystal and the Q2 output coil and capacitor values to suit the frequency of choice. The output coil is 59 primary and 12 secondary windings on a T68-2 toroid core. You will need in the order of 79 pF to resonate this coil, thus a 68 pF plus a 5-20 pF variable works well. If you do not have a T68-2 core, a T50-2 could be used with 30 AWG wire which is a bit difficult so you could opt for an XL of 200 ohms which would be 40 primary and 8 secondary turns on a T50-2 core. You would need ~200 pF to resonate such a coil and the BFO would have reduced output power, but would still work okay. If you design your own tank, you probably should not use an XL/XC less than 190 ohms for the output stage at 4 MHz. Use a 5:1 turns ratio for primary to secondary windings.



It



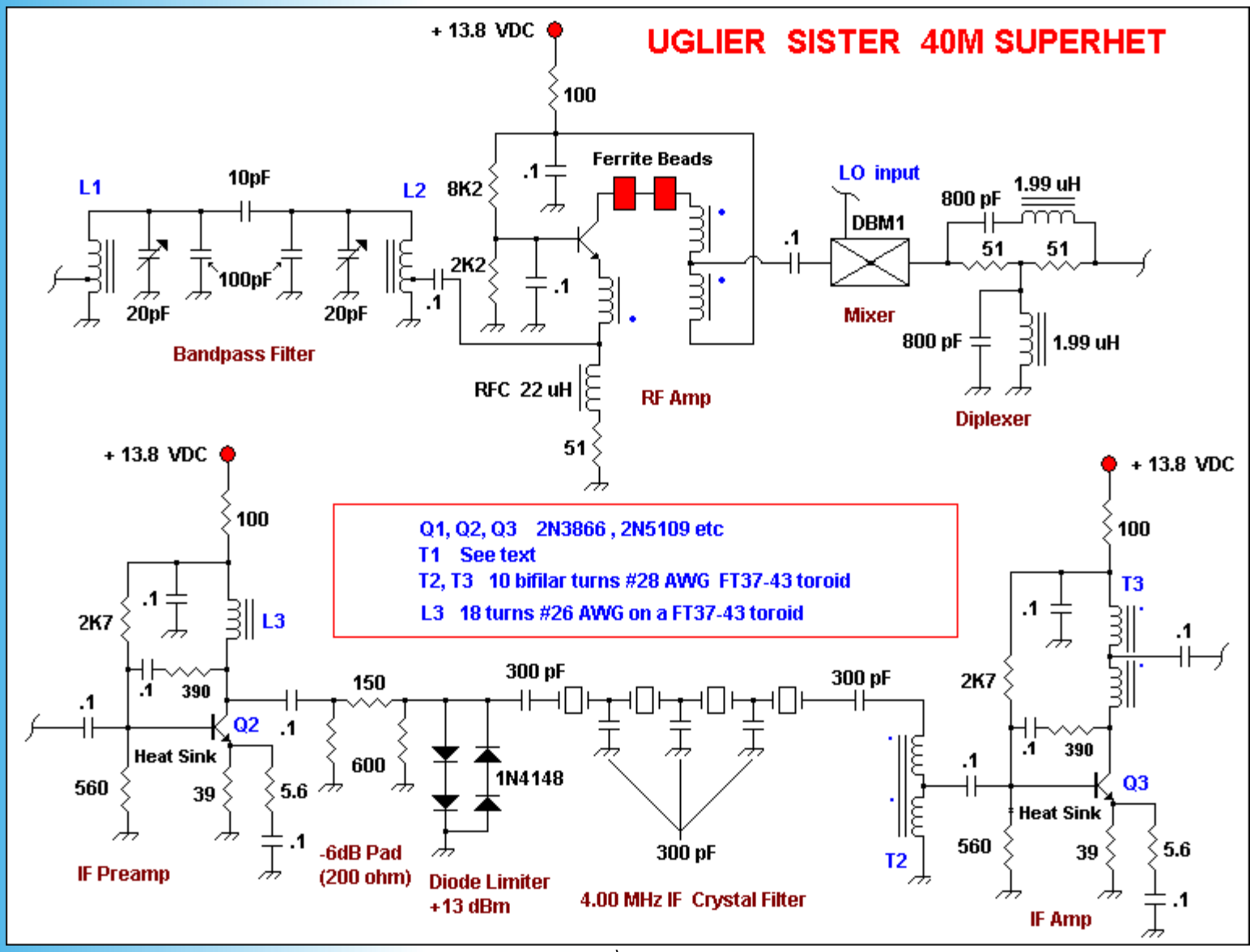
wouldn't hurt to follow the Popcorn Superhet BFO with a low pass filter. A simple pi-type 3 element filter is suitable. Since I had a couple of 470 pF caps left over, I made a low pass filter using a 6.8 MHz cutoff frequency with a 1.17 uH inductor. A 1.17 uH inductor using a T37-2 core requires 17 turns while on a T50-2 core requires 15 turns. Caps were inexpensive monolithic ceramics from Digi-Key. The filter is shown below, L1 = 1.17uH , C1 and C2 = 470 pF. If you want, you may use other cutoff frequencies to suit any junk box caps you have on hand and use an XL/XC of 50 ohms.

Construction

Keep component leads as short as possible to promote stability. Do not omit the zener diode D1. I thought of using a small ferrite bead on the base of Q2 and will try it on the next version of this BFO that I make. The BFO, if unstable can break into all sorts of weird AF sound effects that are very annoying. The BFO should ideally be in a shielded box, but many amateurs have good results by just placing the BFO strategically in the receiver chassis. When winding L1, make sure you wind the secondary coil so that the secondary's grounded end is on the cold side of the

primary coil that connects to the 100 ohm resistor and positive voltage. I just pick one end of the primary coil and close wind the secondary coil from this point in the same direction for the correct number of windings. Then the starting point of the secondary can be grounded and the other end connected to the detector or a low pass filter. After building the coil, temporarily place a 51 ohm resistor from the secondary output lead to ground. Then you can test and peak the tuning of the output tank into a load. Leave the resistor in place for all your testing work such as crystal matching or RF stage testing until you are ready to connect it to the product detector. The 60 pF variable cap connected between the crystal and ground is used to set the beat frequency of the BFO. Adjust it to get a good sounding beat note when listening to a station. You will notice that the beat frequency can be tuned roughly by just listening to 40 meter band noise in the headphones and then tweaked on an active QSO.

QRPHB Archive Schematics



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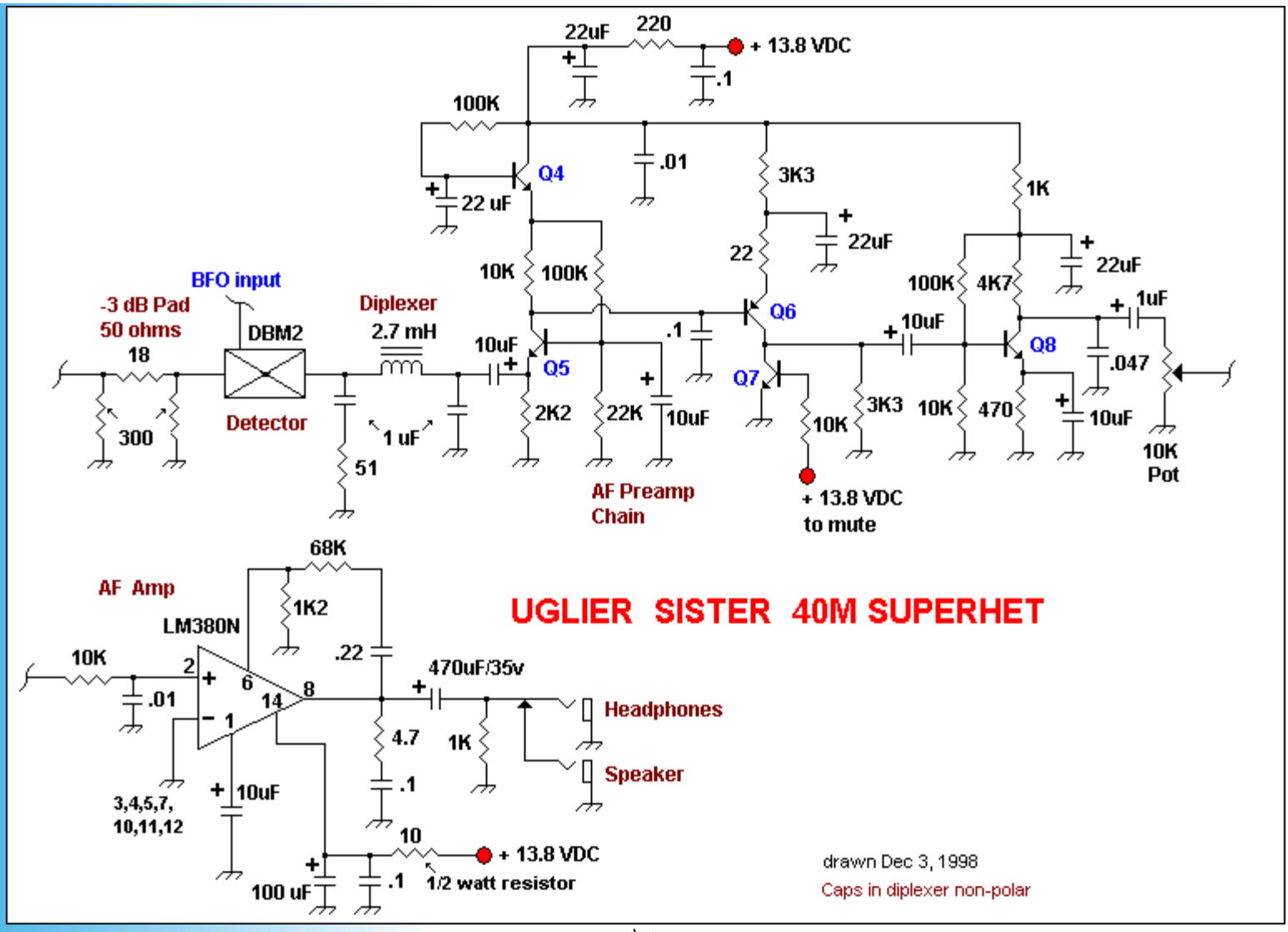

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Uglier Sister Superhet Receiver

This is a 40 M superhet that I built a few years ago. While I was happy with this receiver, it has been discontinued from the main listings on the QRPHB website. I have had a couple requests for this schematic, so I have created this archival web page.
I stood a lot of current in the IF preamp and main IF amplifiers, which is not really necessary, particularly in the case of the post-filter amp. Of course raising the emitter resistor values is easy to do in these amps if you do wish to lower the current. There also is no audio filtering in the AF preamp stages, which is something that I often use in better quality receivers that I build now.

VFO and BFO

VFO schematic
BFO schematic

Construction Information

Refer to this web page for details on the xtal filter, diplexer and construction information.

The Norton RF preamp is discussed on the RF Preamp web page.

Return to the QRP HomeBuilder Homepage

Diplexer Supplemental Page

More Information on the Bridge-Tee Diplexer

Introduction

No web page on the QRPHB web site has furnished more email and controversy than the Diplexer Web page. This page has received much work to improve its content and accuracy and the email received has been greatly appreciated. Presented is the detailed math and some information regarding the Bridge-Tee Diplexer by Helmut Strickner, VK4STR and Wes Hayward, W7ZOI.

VK4STR

The T-bridge diplexer math calculation (also known as the W1JR diplexer)

The formulae presented and used in the program for the Bridge-T Diplexer uses a Q of 1. Using a Q = 1 resonator is not optimal because we don't want a "broad peaking response" of the series LC resonator in the T-bridge diplexer. Our goal should be to only let the desired IF frequency pass and attenuate all other frequencies as much as possible by properly terminating them into 50 Ohms in the diplexer.

The optimal Q for HF frequencies is around 10, less for higher frequencies. The higher Q of the series resonator effectively increases the attenuation of undesired harmonic frequency components from the mixer IF-port, and the reflected products from the crystal back to the IF-port. The IM performance (IP3) of the receiver is improved.

The possible improvement in IP3 depends mainly on the following stages. The post-mixer amplifier following the diplexer has to handle huge impedance variations at the input of the crystal filter. Crystals do not absorb but reflect any products they cannot handle. These products are then reflected back to the IF-port of the mixer deteriorating the IM performance.

A low Q diplexer lets these reflections from the crystal pass unhindered and most mixers will react with degraded intermodulation performance. It is also important to note that a good termination insensitive post-mixer amplifier should follow the diplexer.

The best results for a high intercept (IP3) receiver can be obtained by choosing:

1. High level mixer
2. T-bridge diplexer with optimised Q (Q=10 for HF)
3. Termination insensitive post-mixer amplifier (see N6NWP - QEX July 1995)

I believe that the intermodulation performance of many receivers can be improved by some simple measures like a properly designed diplexer. The T-bridge diplexer from W1JR is the best one around, go for it.

A brief example of the calculation for a 10.7MHz diplexer follows in Figure 1. The value of K should be 10 for short wave and 2 to 4 for VHF (2 meter band). K is the ratio of L to C.

I hope that you can use this example to calculate the values for other diplexer frequencies.

To fully understand the basis of this calculation please refer to Meinke/Grundlach: "Taschenbuch der Hochfrequenztechnik", Springer-

$$K = \frac{1}{Z}\cdot\sqrt{\frac{L}{C}} \quad (K = 10) \qquad \omega_r = 67\,196\,000$$

$$L_1 = \frac{K\cdot Z}{\omega_r} = \frac{10\cdot 50}{2\cdot 3{,}14\cdot 10{,}7\cdot 10^6} =$$
$$= 7{,}44\cdot 10^{-6} \approx 7{,}5\ \mu H$$

$$C_1 = \frac{1}{\omega^2\cdot L_1} = \frac{1}{(2\cdot 3{,}14\cdot 10{,}7\cdot 10^6)^2\cdot 7{,}44\cdot 10^{-6}} =$$
$$= 29{,}7\cdot 10^{-12} \approx 30\ pf$$

$$L_2 = C_1\cdot Z^2 = 29{,}7\cdot 10^{-12}\cdot 50^2 =$$
$$= 7{,}42\cdot 10^{-8} \approx 74\ nH$$

$$C_2 = \frac{L_1}{Z^2} = \frac{7{,}44\cdot 10^{-6}}{50^2} =$$
$$= 2{,}97\cdot 10^{-9} \approx 3\ nF$$

Figure 1

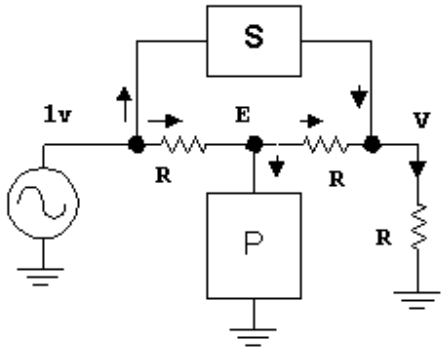
73!
Helmut VK4STR

Verlag Berlin/Heidelberg. Page 127.

W7ZOI

Attached are some outputs from Mathcad, Version 7.0. The "proper" equations are derived, along with some other results. Feel free to put these on the web, although this is mathematics rather than the usual "formulas" and there might not be any interest in such stuff.

As it turns out, the formula that Todd presented in completely accurate for the case of Q=1. That is, both the series and parallel tuned circuits have inductors and capacitors with reactance of 50 Ohms. Evaluation then shows that S11=0 for all frequency, while S21 peaks broadly at the resonant frequency. If you pick a higher Q, the inductor in the series tuned circuit increases by Q and the capacitor in the parallel one increases by Q. Resonance is maintained in both. Again, you then get a perfect match at all frequencies and a S21 peak at resonance. But I don't know if this is what Joe had in his original work.



Shown is a general bridge T type network with series impedance S and parallel impedance P. We wish to find the input impedance seen by the generator. But to do this, we need to know the input current. The input voltage is set at 1, so Zin=1/Iin. To calculate this, we must know E and V. These are obtained with a nodal analysis.

We write the nodal equation at E :

$$\frac{(1-E)}{R}=\frac{E}{P}+\frac{(E-V)}{R}$$

We solve this for E:

$$\frac{(1-E)}{R}=\frac{E}{P}+\frac{(E-V)}{R}\ \text{solve},E\ \rightarrow\ \frac{-\left(\frac{1}{R}+\frac{1}{R}\cdot V\right)}{\left(\frac{-2}{R}-\frac{1}{P}\right)}$$

We now write the equation at V and solve it for E:

$$\frac{(1-V)}{S}+\frac{(E-V)}{R}=\frac{V}{R}\ \text{solve},E\ \rightarrow\ -\left(\frac{1}{S}-\frac{1}{S}\cdot V-\frac{2}{R}\cdot V\right)\cdot R$$

We now equate the two expressions for E to each other and solve that for V.

$$-\left(\frac{1}{S}-\frac{1}{S}\cdot V-\frac{2}{R}\cdot V\right)\cdot R=\frac{-\left(\frac{1}{R}+\frac{1}{R}\cdot V\right)}{\left(\frac{-2}{R}-\frac{1}{P}\right)}\ \text{solve},V\ \rightarrow\ \frac{-\left[\frac{-R}{S}+\frac{1}{\left[\left(\frac{-2}{R}-\frac{1}{P}\right)\cdot R\right]}\right]}{\left[\frac{R}{S}+2+\frac{1}{\left[\left(\frac{-2}{R}-\frac{1}{P}\right)\cdot R\right]}\right]}$$

which simplifies to $V=\dfrac{\left(2\cdot R\cdot P+R^2+P\cdot S\right)}{\left(2\cdot R\cdot P+R^2+3\cdot P\cdot S+2\cdot S\cdot R\right)}$

We now take one of the expressions for E and put V in it to obtain a complete solution.

$$E=\ 2\cdot P\cdot\frac{(R+S)}{\left(2\cdot R\cdot P+R^2+3\cdot P\cdot S+2\cdot S\cdot R\right)}$$

The next chore is to evaluate the input impedance, Zin. We get this by calculating the input current. The input voltage is set at 1, so Zin=1/ i-in.

$$\text{i in}=\ \frac{(1-V)}{S}+\frac{(1-E)}{R}\qquad \text{So } Z=\ \frac{1}{\left[\frac{(1-V)}{S}+\frac{(1-E)}{R}\right]}$$

which simplifies to $Z\text{ in}=\dfrac{-1}{(-R+R\cdot V-S+S\cdot E)}\cdot S\cdot R$

But to get a meaningful expression, we must sub V and E into this. When we do so, we get

Figure 2

$$Z\,in = \frac{\left(2\cdot R\cdot P + R^2 + 3\cdot P\cdot S + 2\cdot S\cdot R\right)}{\left(2\cdot R\cdot P + 3\cdot R^2 + P\cdot S + 2\cdot S\cdot R\right)}\cdot R$$

This is now a useful result. We can, for example, let S, P, and R be resistor values. We can then design bridge T attenuators. But let's continue with the analysis. Let's ask the question "What are the conditions that will provide an input impedance of R?" We merely set the Zin expression equal to R, and then solve that equation for S or P. We will solve for P.

$$\frac{\left(2\cdot R\cdot P + R^2 + 3\cdot P\cdot S + 2\cdot S\cdot R\right)}{\left(2\cdot R\cdot P + 3\cdot R^2 + P\cdot S + 2\cdot S\cdot R\right)}\cdot R = R \text{ solve, } P \rightarrow \frac{R^2}{S}$$

Hey, that's a cool result, and it's even rather simple. This is an expression for P in terms of S. If we use it for attenuators, it says that our pad is matched with these conditions. We can combine these results with the equation for V to get the attenuation. A couple of interesting matched pads are shown:

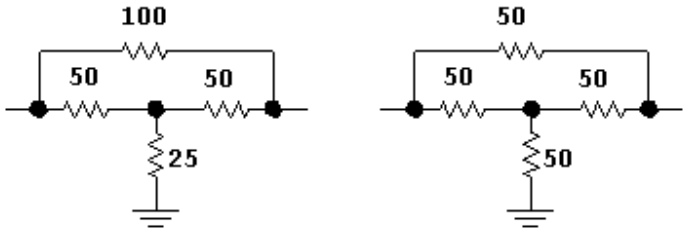


Figure 3

The expression for V was $\frac{\left(2\cdot R\cdot P + R^2 + P\cdot S\right)}{\left(2\cdot R\cdot P + R^2 + 3\cdot P\cdot S + 2\cdot S\cdot R\right)}$ But the conditions for a match were $P = \frac{R^2}{S}$ So, we put this in for P.

$$\frac{\left(2\cdot R\cdot P + R^2 + P\cdot S\right)}{\left(2\cdot R\cdot P + R^2 + 3\cdot P\cdot S + 2\cdot S\cdot R\right)} \text{ substitute, } P = \frac{R^2}{S} \rightarrow \frac{\left(2\cdot\frac{R^3}{S} + 2\cdot R^2\right)}{\left(2\cdot\frac{R^3}{S} + 4\cdot R^2 + 2\cdot S\cdot R\right)}$$

which simplifies to $V = \frac{R}{(R+S)}$ So the two pads have voltage outputs of 0.5 and 0.333 and the resulting attenuations are 6 and 9.54 dB.

But enough about the attenuators. Let's get back to the original question, that of the diplexer. The circuit of interest is one where the series network is a series tuned circuit while the parallel one is a parallel tuned circuit. But we want to know the conditions that will generate a perfect match. With such a circuit, it is clear that a match will exist at resonance so long as both circuits are resonant at the same frequency. The impedance of the series tuned circuit goes to zero while that of the parallel TC becomes infinite (neglecting loss.)

We write the susceptance (imaginary part of admittance) directly for the parallel tuned circuit as

$B_p = \omega \cdot C_p - \frac{1}{\omega \cdot L_p}$ But the reactance is the recipricol of this, $\frac{1}{\left(\omega \cdot C_p - \frac{1}{\omega \cdot L_p}\right)}$

which simplifies to become $\frac{1}{\left(\omega^2 \cdot L_p \cdot C_p - 1\right)} \cdot \omega \cdot L_p$

We know that this is resonant at the same frequency as the series tuned circuit with L and C.

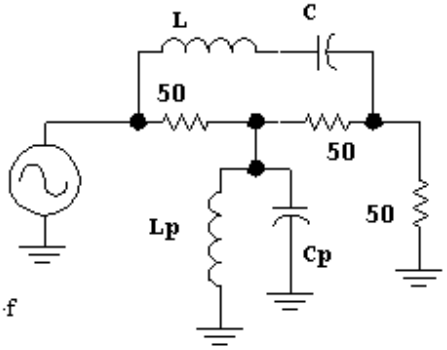
So, $L \cdot C = L_p \cdot C_p$ and $L_p = \frac{L \cdot C}{C_p}$ We substitute these in the above expression:

$X_p = \frac{1}{\left[\omega^2 \cdot (L \cdot C) - 1\right]} \cdot \omega \cdot \frac{L \cdot C}{C_p}$ But earlier analysis showed us that this reactance was related to the series reactance:

$P = \frac{R^2}{S}$ with $S = \omega \cdot L - \frac{1}{\omega \cdot C}$ This lets us set up an equation that is then solved for Cp.

$\frac{1}{\left[\omega^2 \cdot (L \cdot C) - 1\right]} \cdot \omega \cdot \frac{L \cdot C}{C_p} = \frac{R^2}{\left(\omega \cdot L - \frac{1}{\omega \cdot C}\right)}$ solve, $C_p \rightarrow \frac{L}{R^2}$

This is a particularly simple, yet useful result



1) Pick a crossover frequency f, Hz. Then, $\omega = 2 \cdot \pi \cdot f$

2) Pick a series inductor L. Use one with a reactance of around 50 Ohms if you want a very broad response. Make it larger for narrower bandwidth.

3) Resonate the series L with a cap, $C = \frac{1}{\omega^2 \cdot L}$

4) Pick a parallel cap $C_p = \frac{L}{R^2}$

5) Resonate that cap with a parallel inductor $L_p = \frac{1}{\omega^2 \cdot C_p}$

Figure 4

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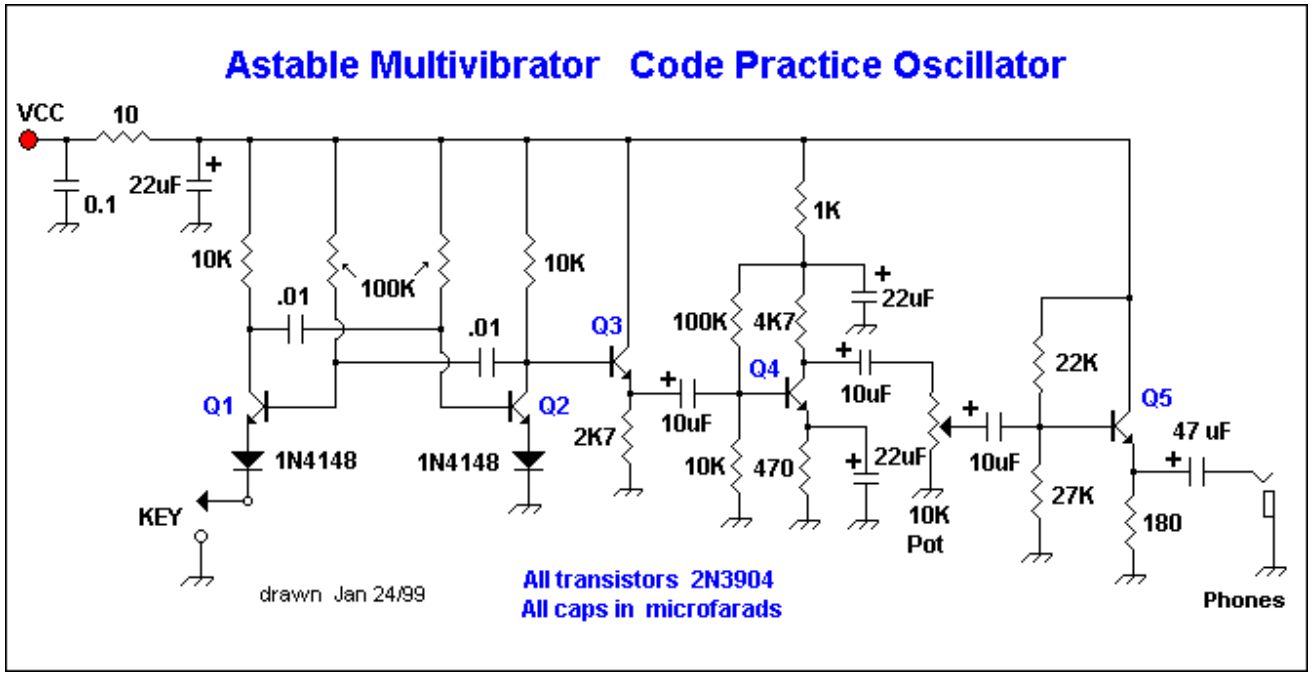
Astable Multivibrator Projects

Discussion:

Astable or free-running multivibrators have been used in home-built amateur radio equipment for many years. The basic circuit is a two stage amplifier with AC-coupled feedback from output to input. One transistor stage is on (conducting current) while the other is off (not conducting current) until the stages switch conducting states repeatedly at a specific frequency. The oscillation frequency is set by the resistor and capacitor values connected to the base terminal of each stage. This RC network determines how long the transistor stays in the off position.

Presented are two projects which utilize astable multivibrators built using the ubiquitous 2N3904 BJT. The first project is a code practice audio-frequency oscillator while the second is a simple , no-frills electronic keyer for keying a transmitter. Either circuit would be a great first project to learn how to build circuits using Ugly Construction.

Code Practice Oscillator



Above is the schematic for a simple Morse code practice oscillator. This circuit was originally built with 2N3904 transistors, however many different NPN transistors could be substituted as required.

Tracing the circuit from left to right first brings us to the multivibrator circuit which is composed of Q1 and Q2. The oscillation frequency of the multivibrator is ~ 700 hertz and is set by the RC network formed by the 100K resistor and the 0.01 uF capacitor connected to each transistor base terminal. The approximate time off for each transistor is given by the following formula:

[Time 0ff = 0.7 * R * C] with R in ohms and C in farads.

It maybe more practical to leave the resistance value fixed and vary the capacitor value to obtain a desired oscillation frequency. Rearranging the above formula allows this :

[Total Time Off = 1 / Frequency] with Total Time Off being the total number of seconds that both transistors are off and Frequency is in hertz.

Once the total time off is known, you must divide that answer by 2 as each transistor is off half of the total time off in this symmetrical circuit. Then you simply solve for the capacitor value:

[Time Half = Total Time Off / 2]
[Capacitor = Time Half / (0.7 * R)] with Capacitor answer in farads.

Lets run the numbers to solve for the capacitor values in the schematic;
R = 100K, desired frequency = 700 hertz.
Total Time Off = 1 / F ----> 1 / 700 = 0.00143 seconds.
Time Half = Total Time Off / 2 ----> 0.00143 / 2 = 0.00071 seconds.
Capacitor = Time Half / (0.7 * R) ----> 0.00071 / (0.7 * 100000) = 0.0000001 farads = 0.01 microfarads.

For 600 hertz, the capacitors would be 0.012 uF and for 400 hertz, 0.018 uF.

As you can see it is maybe necessary to adjust the base resistor value to achieve a specific oscillation frequency. The rule of thumb is that the base resistor should be ~ten times the value of the collector resistor assuming the base-bias resistor is connected to VCC as shown.

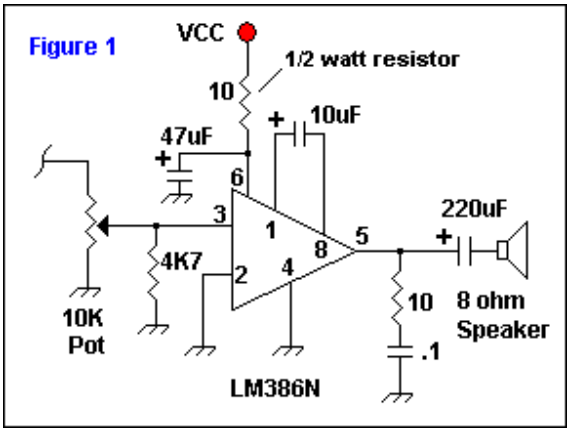
The output of the multivibrator is buffered by the high input impedance of an emitter follower Q3. This serves to prevent the oscillation frequency from changing when the output load is changed. The AF stage connected to the emitter-follower is a standard high gain common-emitter amp that has been used in many of the projects on this web site. As the multivibrator is buffered by 2 amplifier stages, good frequency stability is realized and frequency changes are negligible when turning the volume control pot. The final stage is a common-collector amplifier which can drive low impedance headphones with reasonable volume.

Code Practice Oscillator Project Notes

The voltage / time output waveform of the astable multivibrator is largely a square wave which some people find harsh. Many users prefer listening to a sine wave although that is beyond the scope of this web page. The multivibrator shown has real advantages in that, it is both dependable and tolerant with respect to parts substitutions. Keying the oscillator as shown practically guarantees that the multivibrator will start running each time you hit the key.

VCC can be 9 to 13.8 volts DC and the larger the voltage the greater the volume in the headphones. The B+ decoupling circuit at the top left can be omitted for battery operation or if you prefer not to bother with it. The basic multivibrator and emitter-follower circuit can be used in a transceiver as a sidetone for monitoring keying. A series resistor from the emitter-follower maybe necessary to attenuate/match the sidetone to the transceiver's AF amp.

If you desire speaker level output, the Q5 common-collector final can be omitted and the circuit shown below used. This circuit uses the LM386N and provides up to ~ 0.5 watts into an 8 ohm speaker. Connect the Figure 1 circuit as shown to the 10 uF coupling capacitor connected to the collector of Q4. Do not connect the 10 ohm half-watt resistor to the decoupled VCC shown in the schematic. The power supply to the LM386N AF amp should be directly connected to the VCC, not like the common-collector AF stage shown in the schematic to the right.



Simple Electronic Keyer

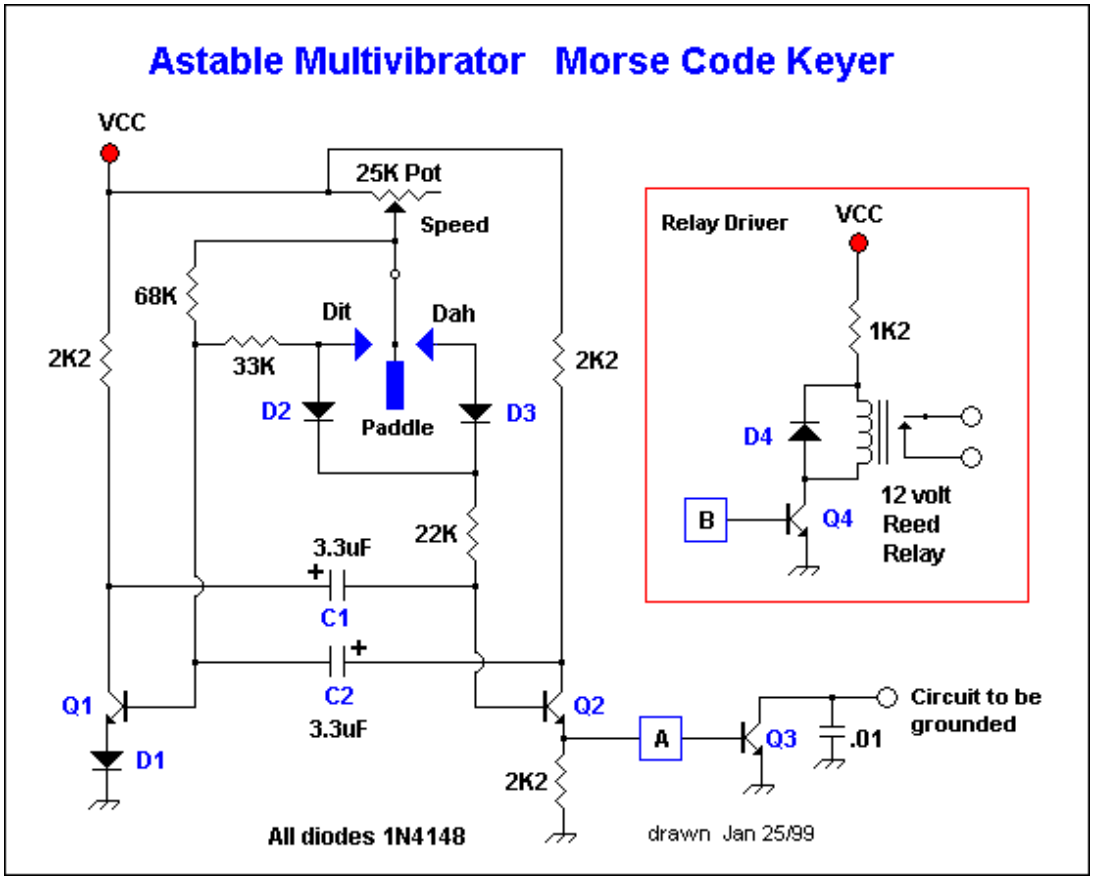
A simple keyer suitable for a popcorn QRP transmitter can be built around an astable multivibrator and an example of such is shown above. The basic design of this keyer is from notes, however the original author of the circuit is unknown. The notes were written in 1973 . I modernized the circuit, added a variable speed control and designed an additional output driver stage.

Keyer Operation

Keying this circuit generates either dits and spaces or dahs and spaces. When the keyer is idle, Q1 is on and Q2 and Q3 are switched off. When the user sends code, Q1 turns off and Q2 and Q3 switch on and in

turn key any device appropriately connected to the collector of Q3. The off-time of Q1 sets the on-time of Q2 and Q3. The off-time of Q2 and Q3 is set by the 22K Q2 base resistor. This off-time is the set time of the spaces and is constant.

The 68K resistor on the base of Q1 is about three times the resistance of the 22K base resistor on Q2 and consequently dahs are ~ three times the length in duration than dits. Spaces and dits are of the same length of time because when sending dits, the 68K base resistor is paralleled with the 33K resistor and effectively the resistance is ~22K ohms. If the optional relay driver transistor Q4 is used instead of Q3, the theory is the same, just substitute Q4 wherever you see Q3.



Keyer Speed

The keyer speed is very sensitive to the power supply voltage and any keyer speeds mentioned are ball-park values. Your results may vary depending on your VCC and component tolerances. If the 25K speed control pot is turned to minimum resistance, the actual power supply voltage will be present on both the paddle common and the top end of the 68K base resistor. This will be the maximum speed for the keyer. In fact, the speed control pot could be omitted if you want to economize and the keyer will run at the maximum speed as determined by C1 and C2. If you do not want the speed control feature, connect the paddle common and the 68K resistor to the VCC supply. Another alternative is to build a two speed keyer by using a switch to switch in or out a fixed resistor to vary the voltage instead of using a potentiometer. A trimmer resistor may also be used for "lid-off" speed adjustments.

Varying the base-bias voltage with a pot changes the charging rate on capacitors C1 and C2. Although, I experienced no problems be careful with some resistances/VCCs as the circuit may be unable to provide enough current to saturate the transistors when the base-bias voltage is at its minimum setting (pot set to maximum resistance). Smaller pots such as 10K can also be used with a more limited ability to reduce the keyer speed below the maximum rate.

To set the maximum rate for the keyer, it is necessary to vary the value of C1 and C2. For this circuit to function correctly, C1 must equal C2. An experiment was conducted with the 25K pot removed and the paddle common and the supply end of the 68K resistor connected to the main B+ terminal. VCC was measured at 13.8 volts. C1 = C2. The words per minute were counted for four different standard capacitance values and the results were as follows:

2.2 uF = 27 WPM
3.3 uF = 23 WPM
4.7 uF = 17 WPM
10 uF = 9 WPM

For this project, I settled with the 3.3 uF value, although personally, I use a 2.2 uF capacitor for C1 and C2. Sending speed can be reduced with the speed control pot or by increasing the time interval between characters and words. With 3.3 uf caps for C1 and C2, turning the 25K speed control pot to maximum resistance dropped the sending rate down to 12 words per minute. If you needed to slow down below 12 WPM, an amateur could send code using the Farnsworth method as mentioned above. At any rate this circuit allows you to determine the maximum speed rate by choosing the C1 and C2 capacitance value to suit your needs.

Q3, Q4 Output Stages

Two different output stages maybe used with this keyer and they will be referred to as the Q3 or Q4 stage.The Q3 stage is a simple transistor switch which will ground any component(s) connected to its collector when turned on during code sending. A variety of transistors maybe used here and care must be taken to ensure that you do not exceed the maximum dissipation of a given BJT.

The Q4 stage is a relay driver. The 12 volt relay used during bench testing was a Radio Shack reed relay that I had in my parts collection. Specifications were SPST 1A @ 125 VAC , 1050 ohms DC coil resistance, part number 275-233. My VCC was 13.8 volts so a 1K2 current limiting resistor was placed between the relay and supply voltage. The relay has suffered no harm despite significant torture but be careful when you are using a VCC greater than 12 volts. This resistance may be dropped down to 470 ohms or omitted if you are using lower voltages. Any relay with a DC coil resistance of 500 to 3000 ohms should work in this circuit.

I have never used this keyer with the Q4 relay driver for QRP operation as I prefer solid-state switching. The Q3 stage in turn can be connected to trigger a PNP transistor switch to supply DC voltage to a keyed transmitter driver and/or to pull a VFO to its offset frequency. A great example of PNP transistor switches can be found in The Ugly Weekender article by KA7EXM and W7ZOI. This article is referenced in the recommended reading list on this web site. Another bonus of the Q3 stage is that it draws less current than the Q4 stage.

Sending Code

If you are used to a deluxe iambic keyer, this keyer will take some getting used to. The dahs in particular can be problematic as they do not self-complete like they do on my station homebrew TTL iambic keyer. After some practice, however most people should be able to send some good morse code with it despite the lack of iambic luxury.

Miscellaneous

When you power up this keyer, it sends a dit. In one version of this keyer, used in a 40M transceiver, I had a 10K pot for the speed control with a built-in switch on the 10K pot to turn the keyer on or off during station setup.

The diodes D1 - D4 can be any common switching diode such as the 1N914. Q1 to Q4 can be any NPN transistor with a Beta greater than 50 such as the 2N3904 or 2N2222. In the test keyer, all BJT's were 2N3904.

For fun, I connected the multivibrator code practice oscillator to the keyer and gave the keyer a workout. They both work great together and are very complimentary. Have fun with astable multivibrators !

Update Oct 2007:

Giovanni, IW7EHC has posted a great web page on this schematic including PCB layout files in FidoCAD format

Check out his excellent Italian and English language web site

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Amateur and Short Wave Radio Electronics Experimenter's Web Site

Supplemental Page for the 10 MHz WWV TRF receiver

I have received a tremendous amount of feedback regarding the 10 MHz WWV TRF receiver. Many builders have constructed their own versions and have improved it. This captures the spirit of the QRP/SWL Homebuilder web site; take an idea and evolve it into something better on the bench.

I will post some of the feedback as well as provide additional comments and images regarding the 10 MHz WWV project on this supplemental web page. Please continue to provide project feedback and help me to improve the QRP/SWL HomeBuilder web site.

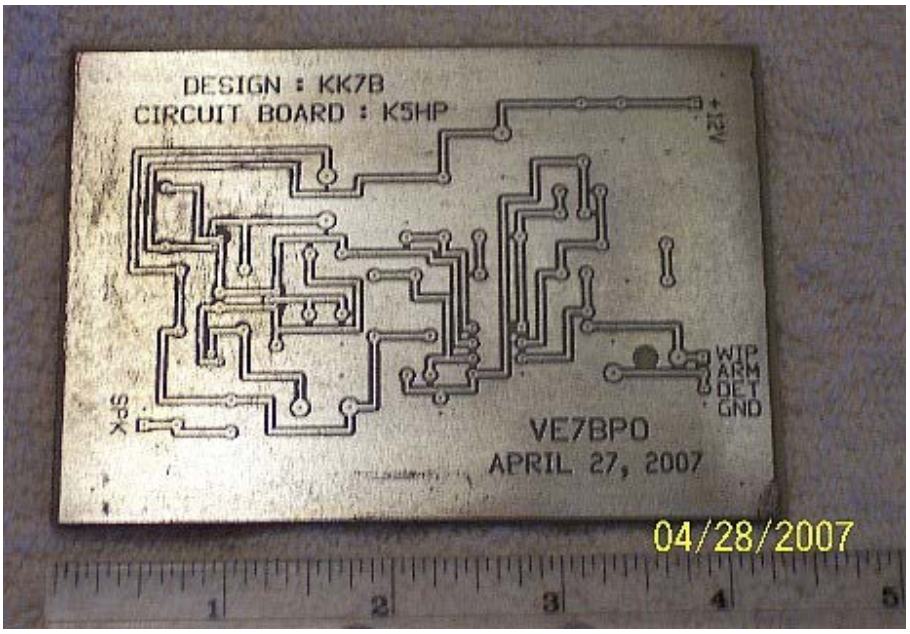
Terry, K5HP
Hi Todd,

I have just finished (few days ago) your design for the 10 MHz WWV Receiver. I reduced you schematic diagrams to CirCad schematics, drug out the rat-net, positioned and connected everything, and etched the boards. I'll include a photo of the 4 finished boards. They are very roomy for extra tinkering. Except for two solder bridges on two very big pads on the tweeker capacitors in the front end, it all worked perfectly first try. Hmmm, I cannot say that for all my projects in the past. After repairing the two shorts to ground of the RF path, the front end then tuned and worked perfectly also.

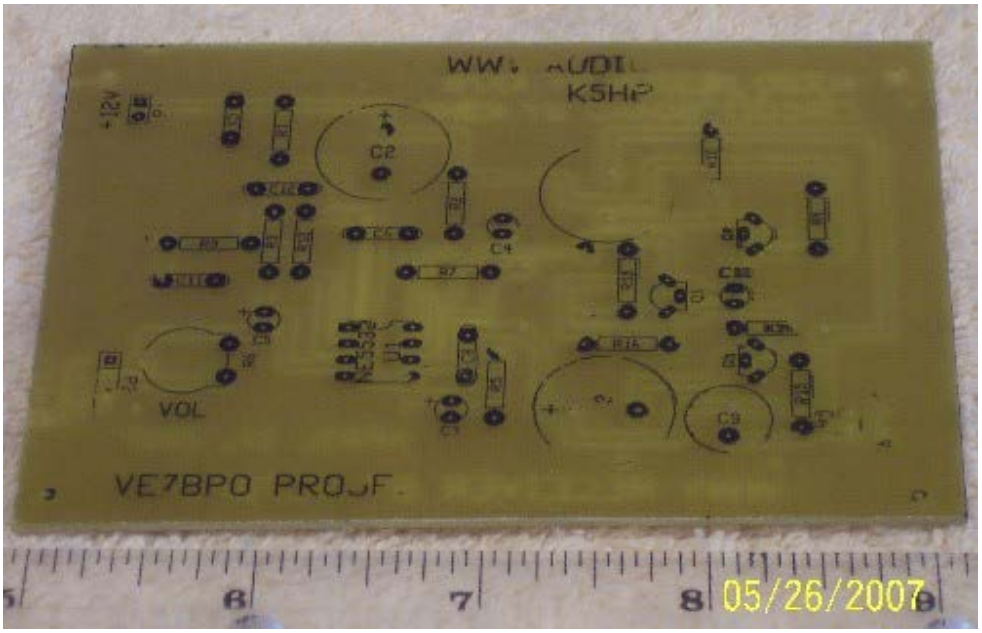
I have not taken the project out to the Lab (barn) and put it on my Motorola Service monitor, but the input to the filter board is roughly 500 uV. It can easily hear WWV at this level with a decent antenna connected to the filter input when WWV is fairly strong at my location. Not bad, considering the mismatch at that location.

Have you or someone else tinkered with an AGC/AVC circuit? At one time I had a nice circuit using an National chip (don't remember the number) which is no longer available that was easy to build and easy to retro-fit into something. It was an AVC which they called an audio AGC.

When I built out the boards I placed proper credit to you and the other developers mentioned on the boards. Should you want my work product it's free for the asking. I'll be glad to send it to you. CirCad in ver 4 is a fairly



Above. Etched Audio amplifier PC board from K5HP.



Above. Audio amplifier PC board overlay for parts placement from K5HP.

complete PCB cad program and it's free too. There is a version 5, but I have not tried it out yet.

My sincere congratulation for producing and error free web article. That is a heck of a more than you can say about many other projects in the past. Most of those in the past had errors a plenty, and some just did not flat work as described. Someone just thought they would as it turned out. Thanks for the fine article and I'll check your site over for more interesting thing to build or learn from. Best wishes and good luck.

Above. A layout of the 4 completed boards by K5HP.

"I use the CirCAD to produce the artwork, then take the laser printer output and bond the plastic toner onto the PCB. Works good most of the time. Sure easier than trying to do photo etch, and silk screening is pretty expensive. The bottom copper is viewed as you would look down onto the top of a circuit board, through to the copper side. The side you are seeing is bonded to the copper surface. Print one out, hold it up to a light source looking through non printed side and it becomes apparent."

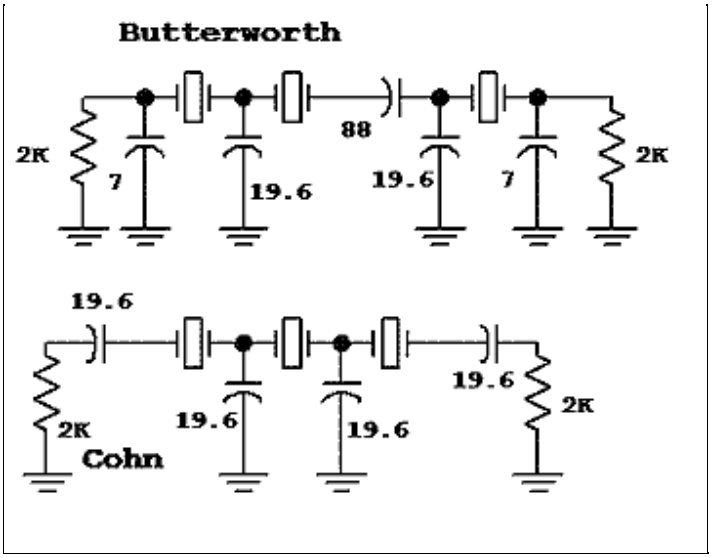
Terry designed and built printed circuit boards. His photographs, etching and overlay files are all bundled together in convenient zipped files which can be downloaded below.

Download the front end zip file

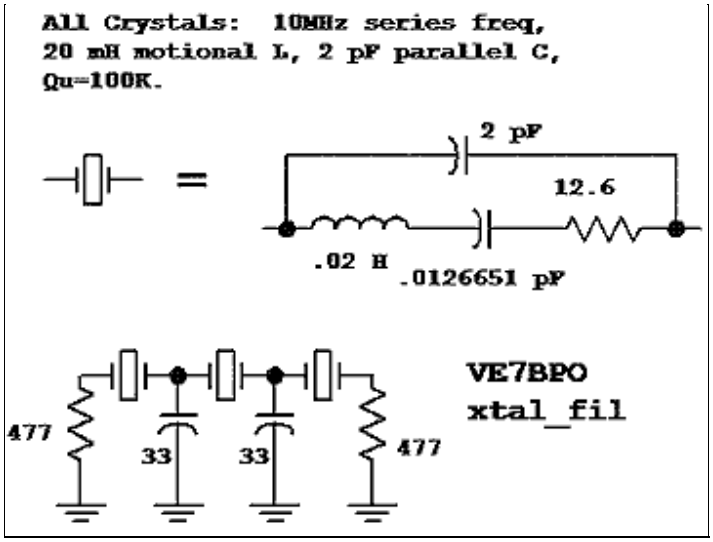
Download the Crystal filter zip file

Download the Detector zip file

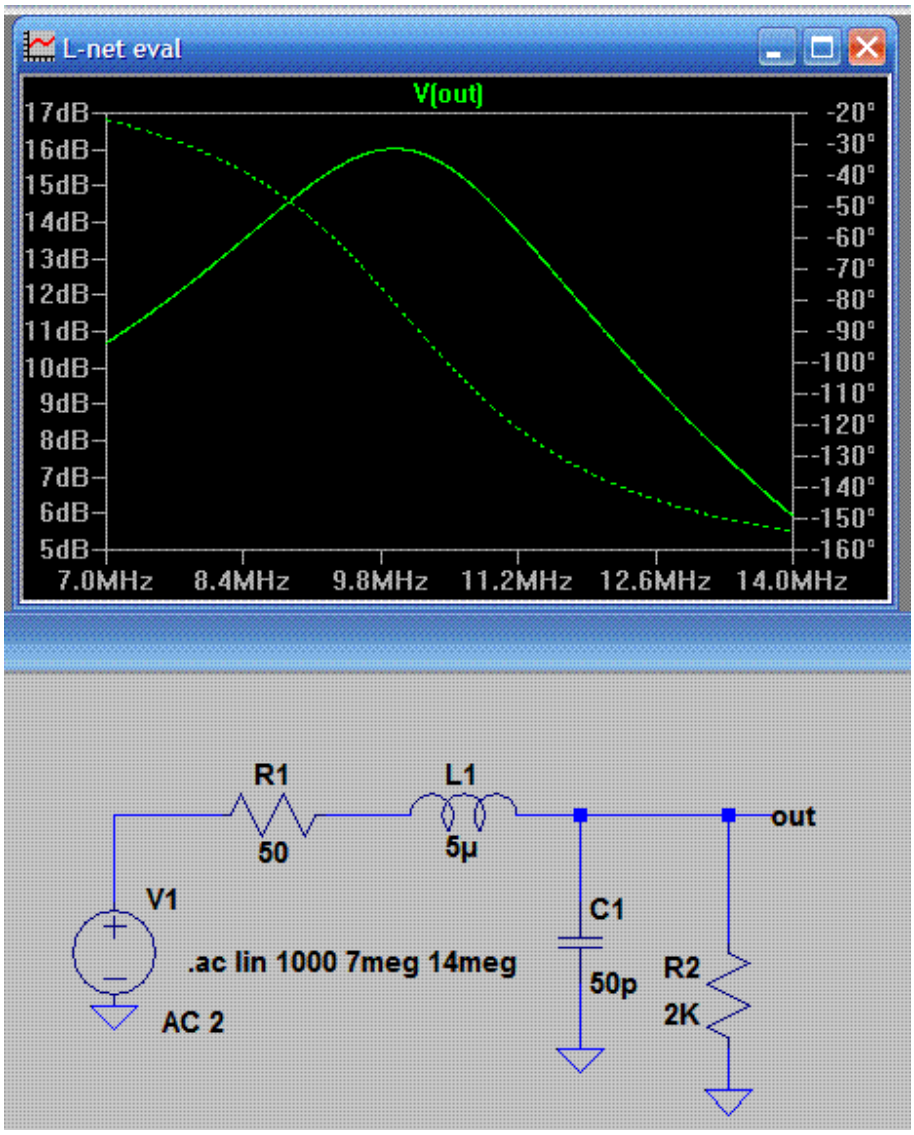
Download the AF amp zip file



Above. Schematics diagrams of 10 MHz crystal filters by Wes, W7ZOI. I did not experiment with the Butterworth type filter.



Above. Schematics diagrams by Wes, W7ZOI.
A schematic representation of the 10 MHz xtal is shown along with the "popcorn" 10 MHz xtal filter I designed.



Above. A simulation of an L network depicted in the L-Match and AM Detector schematic for those interested in driving this detector with a 50 ohm output impedance.

Amateur and Short Wave Radio Electronics Experimenter's Web Site

MF TRF Receiver Supplemental Page

Click here to go to the main project web page This supplemental page is under construction

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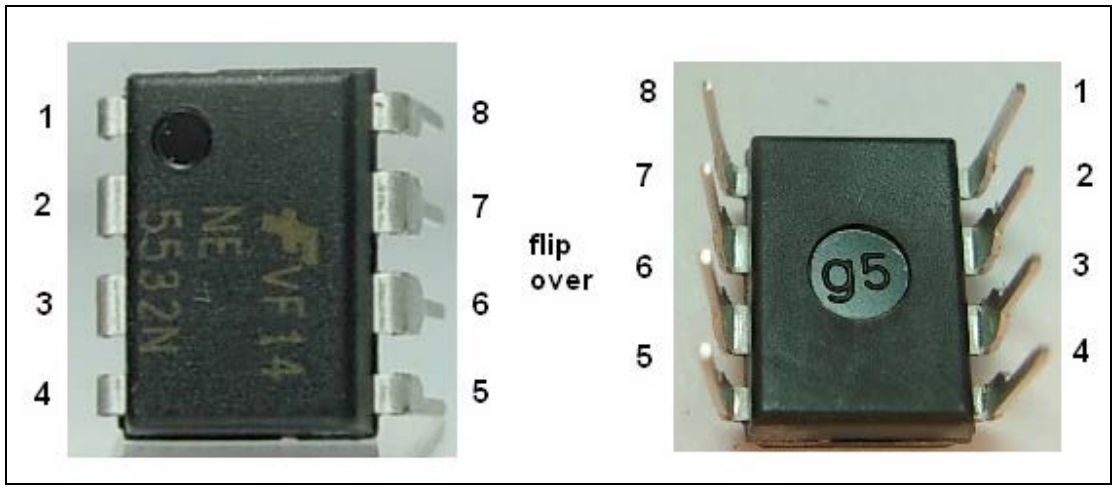
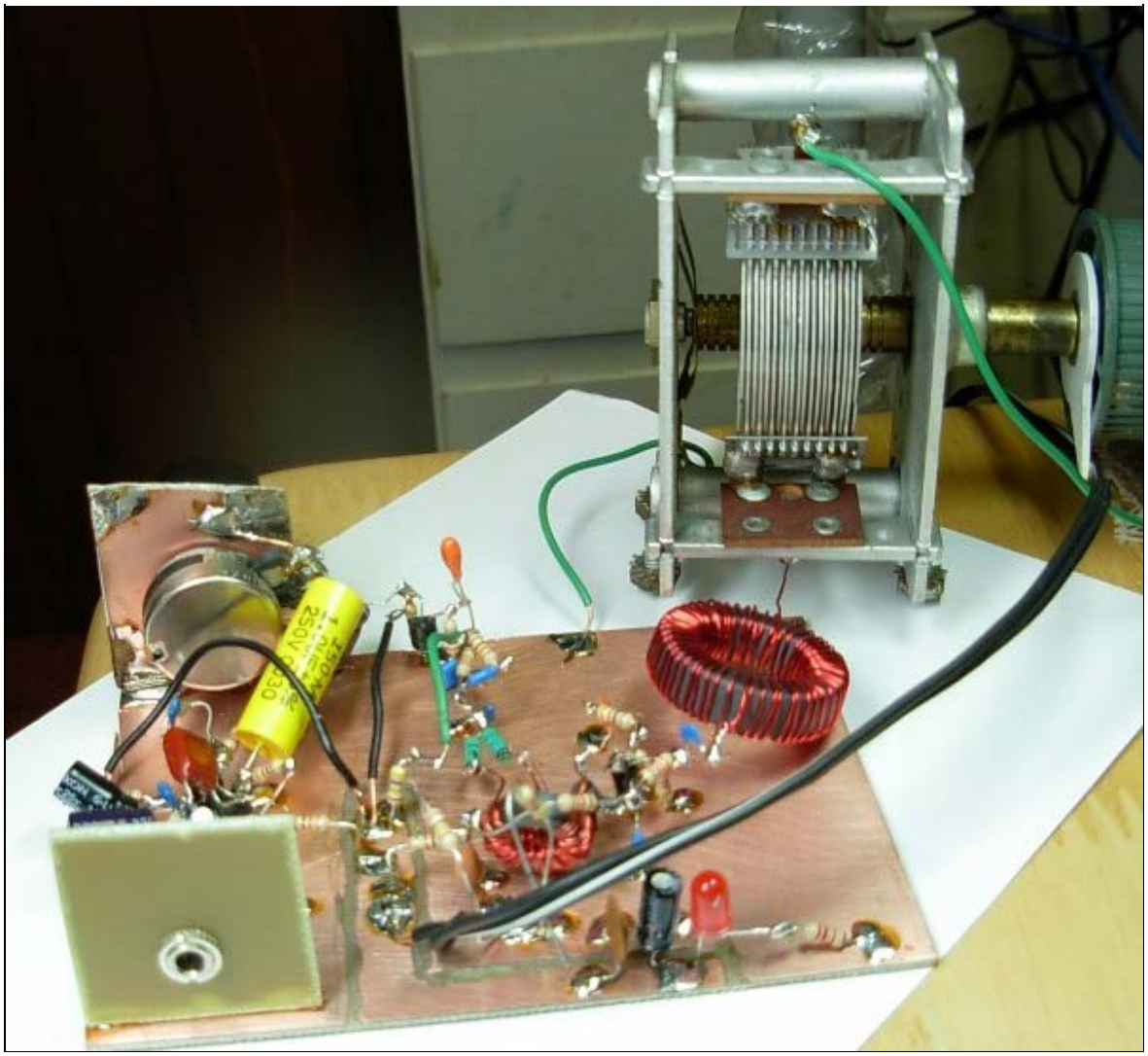
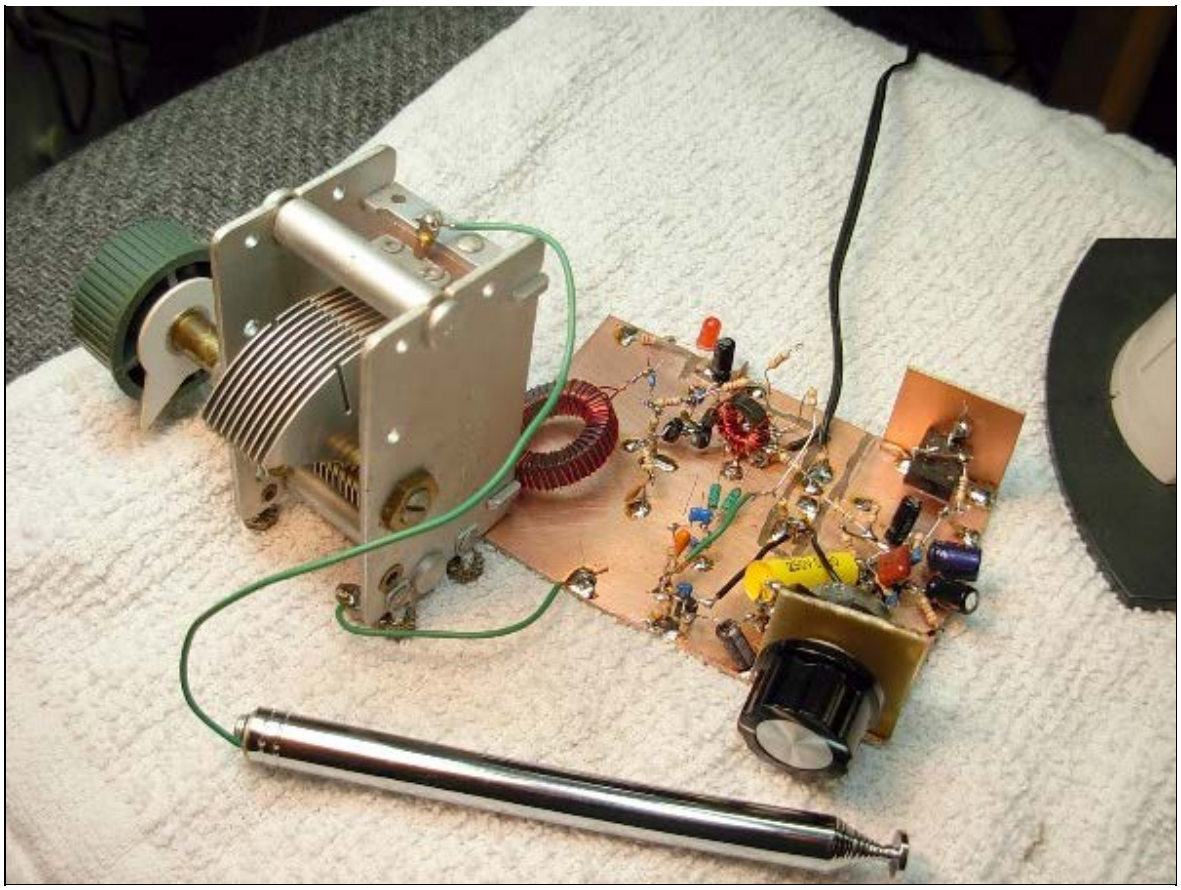
5532 Op amp Pin Out



A Version of the Receiver Using Just L1 and C1 as the band pass filter

You can see a red LED and current limiting resistor in the photos. This is a power on indicator. The telescopic whip antenna can be seen in the lower photo.



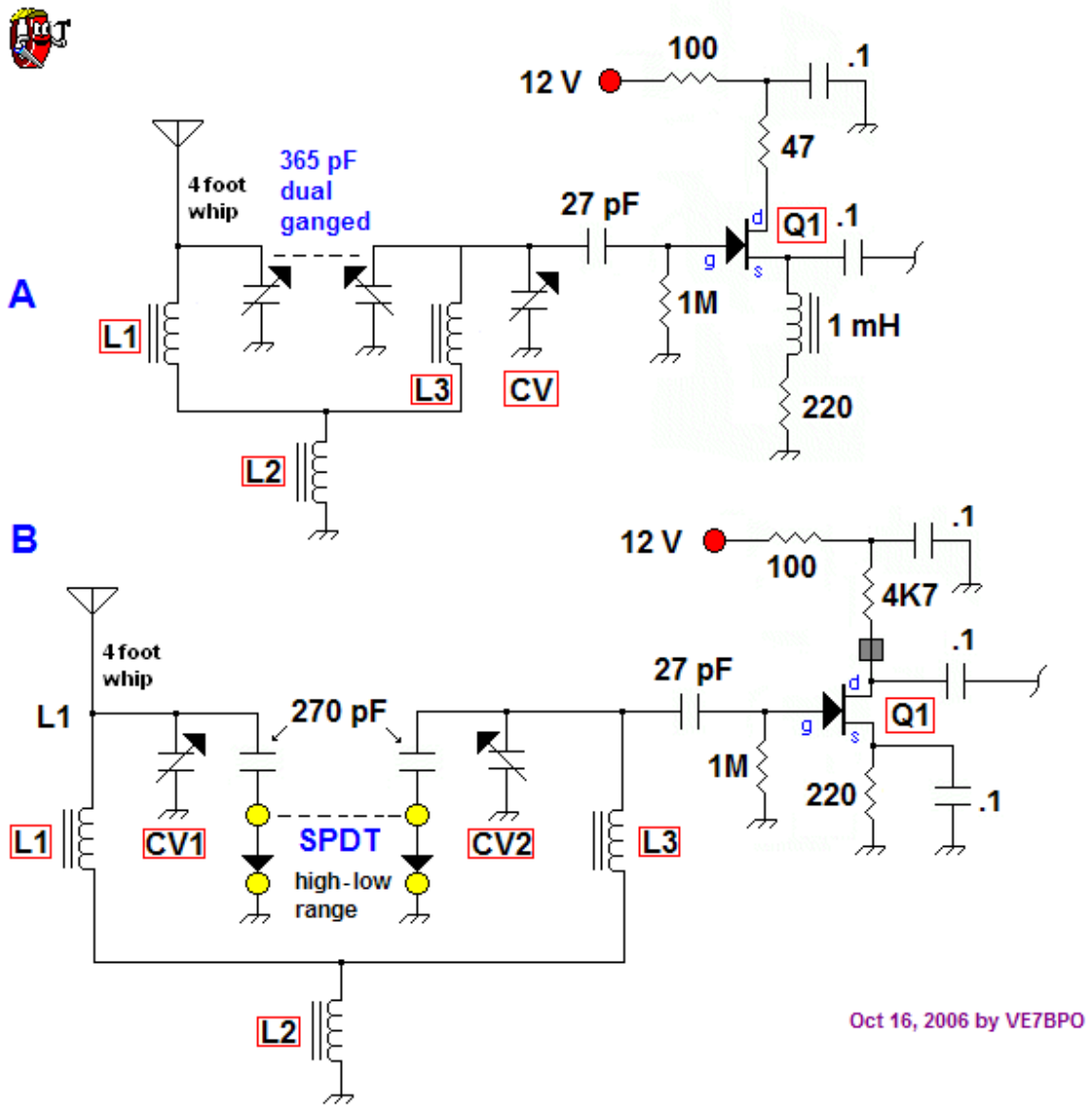


Early Versions of the Project Front End

In section A of the schematic below, a dual-ganged 365 pF air variable capacitor was used for tuning. CV, a small compensating capacitor may be required to adjust for the capacitive differences of the 2 L-C tanks in the front end filter. It may have to be placed on the L1 side of the filter (opposite to what is shown). A JFET source follower was used for Q1 but discarded because stage gain was insufficient with a short whip antenna.

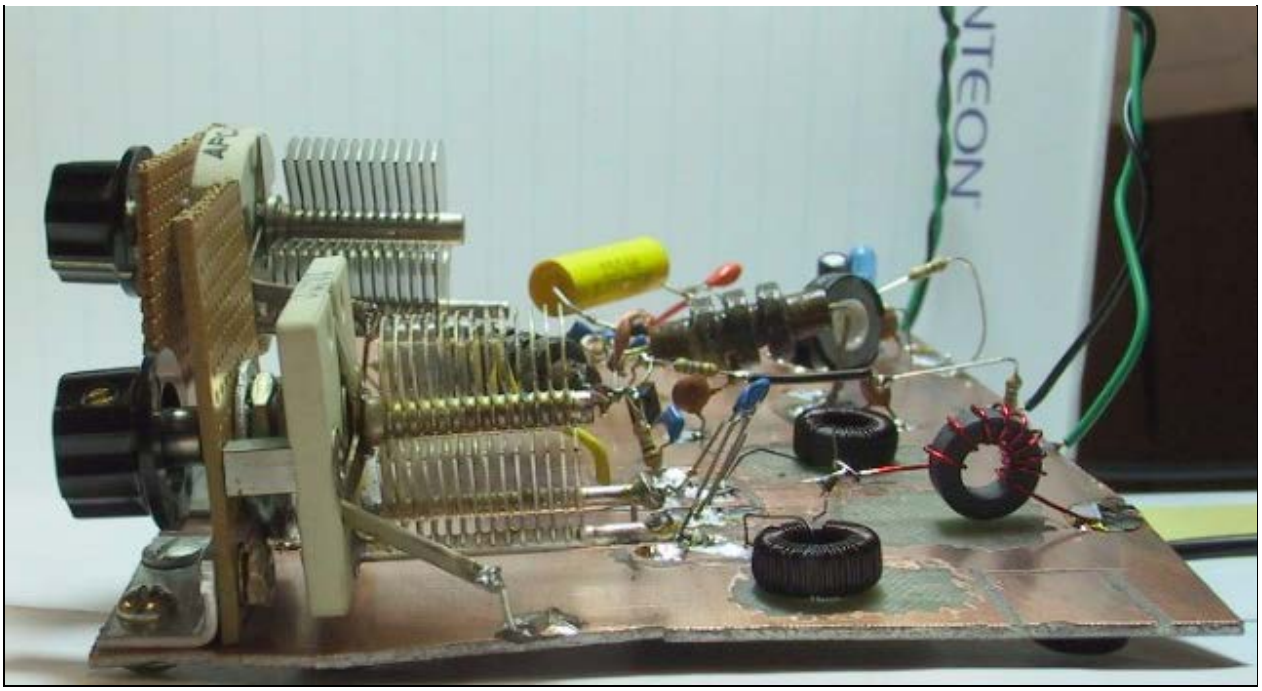
In the version depicted as section B, two 12-110 pF air variable capacitors were used along with switchable 270 pF capacitors to provide 2 ranges of tuning. Although this worked, the switching mechanism was found to degrade selectivity somewhat. The JFET was reconfigured as a common source RF amp. This amp caused some distortion of the RF signal on stronger stations unless the gate resistor was lowered to 47K or less. Ultimately a cascode BJT RF amp was chosen for the project. A photograph of the section A breadboard is shown below the schematic.

Broadcast Radio Band Front End

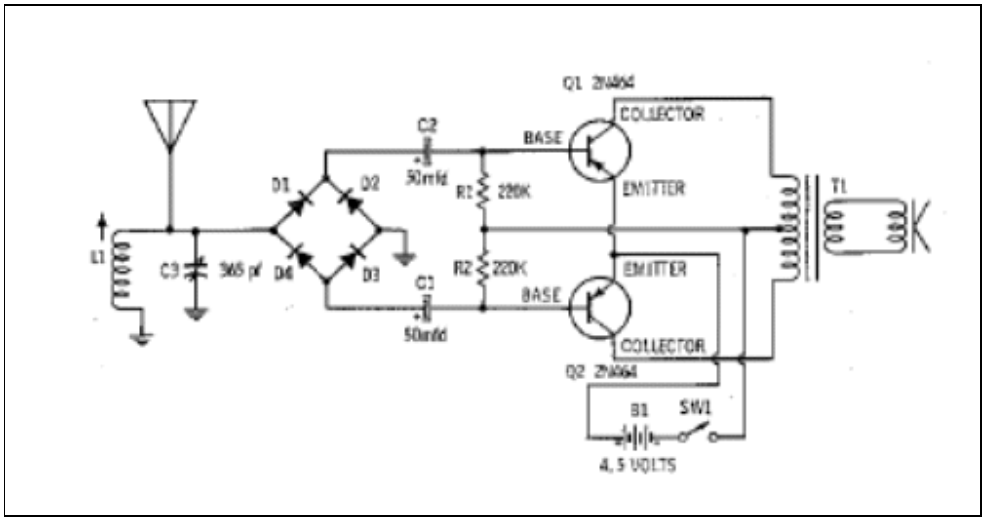


Oct 16, 2006 by VE7BPO

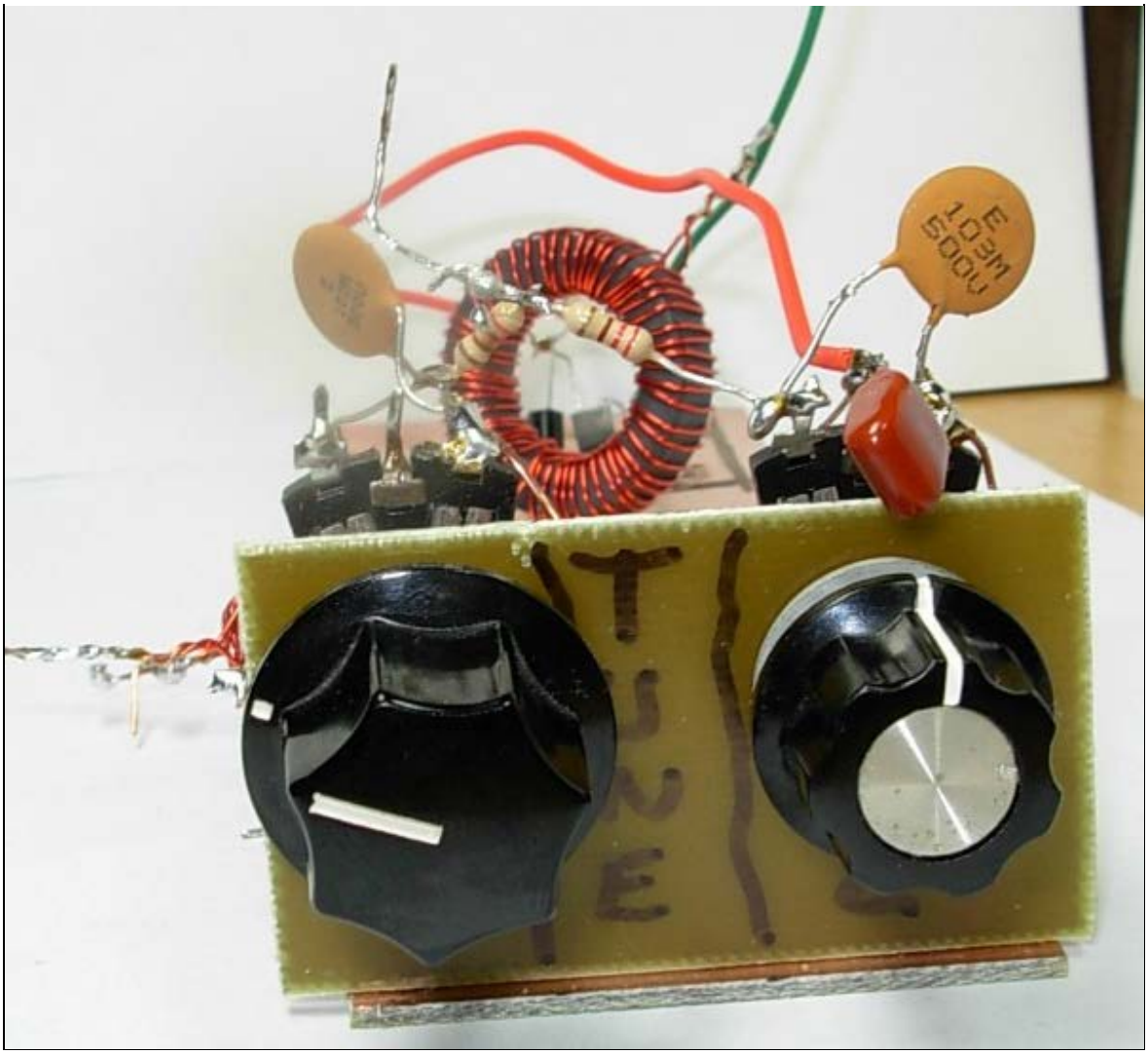
L1 = L3 = 200 uH or 55 turns #28 AWG on a FT50-61 ferrite core
L2 = ~ 5 uH = 9 turns #22 AWG on a FT50-61 ferrite core
CV = small value trimmer capacitor for tracking - 4.5 - 20 pF
CV1 = CV2 = 12 - 110 pF air variable capacitor
Q1 = nJFET; MPF102 or J310



The inspirational detector schematic used for motivation to develop the full wave detector



Varactor Tuned Front-end Band pass Filter

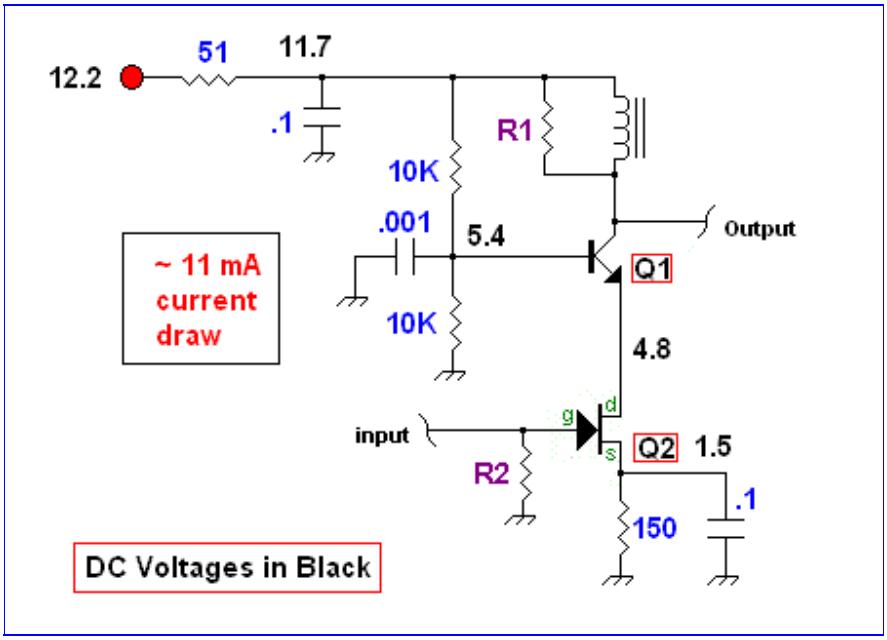


Above photo. The 2 potentiometer varactor controlled band pass filter.

Supplemental to the Cascode Hybrid 5 MHz WWV Receiver

DC Voltages of the Cascode Hybrid Amp

On the right is the basic hybrid cascode amplifier schematic. Note that the Q1 voltage divider bias has two 10K resistors as compared to the 12K and 10K combination used in the final project. A slightly lower Q1 bias voltage was ultimately chosen along with a 150 ohm Q2 source resistor to enhance stability. Early bread boards of this amplifier used a 47 - 100 ohms Q2 source resistor along with the slightly higher Q1 bias, but suffered from parasitic oscillations.



The DC voltages are rounded to 1 decimal point. These are ball park values for reference purposes. The typical stage current draw was around 11 mA. R1 sets the output impedance and is optional. R2 sets the input impedance. This amp is very versatile in that it can be used to match a wide variety of input and output impedances.

The **main 5 MHz WWV receiver web page** which this web page supplements is here.
Click here for a recording of 5 MHz WWV made with the Icom R-75 receiver using the 6 KHz AM filter and the stock speaker. You may wish to compare it to the recorded 5 MHz WWV receiver project audio files linked here , here and here.

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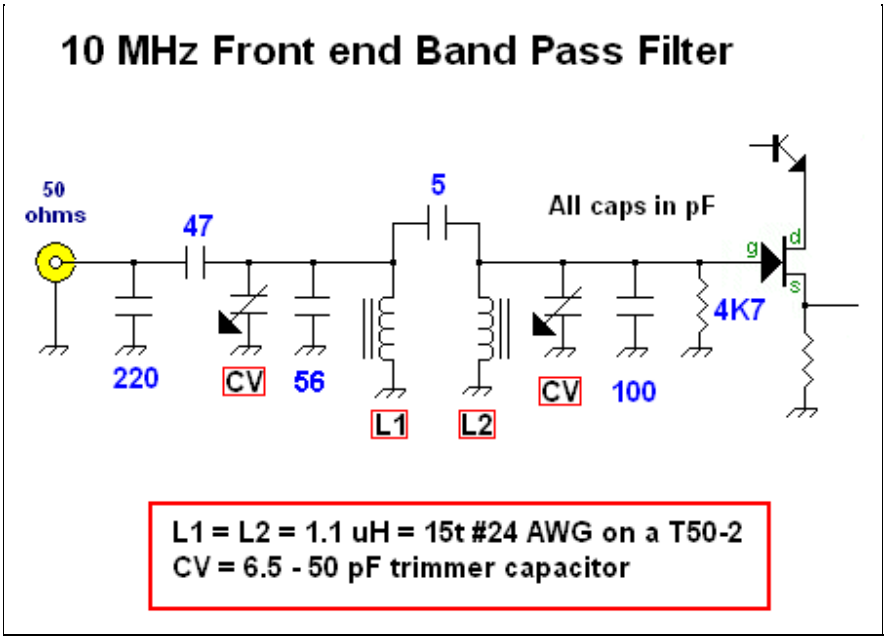
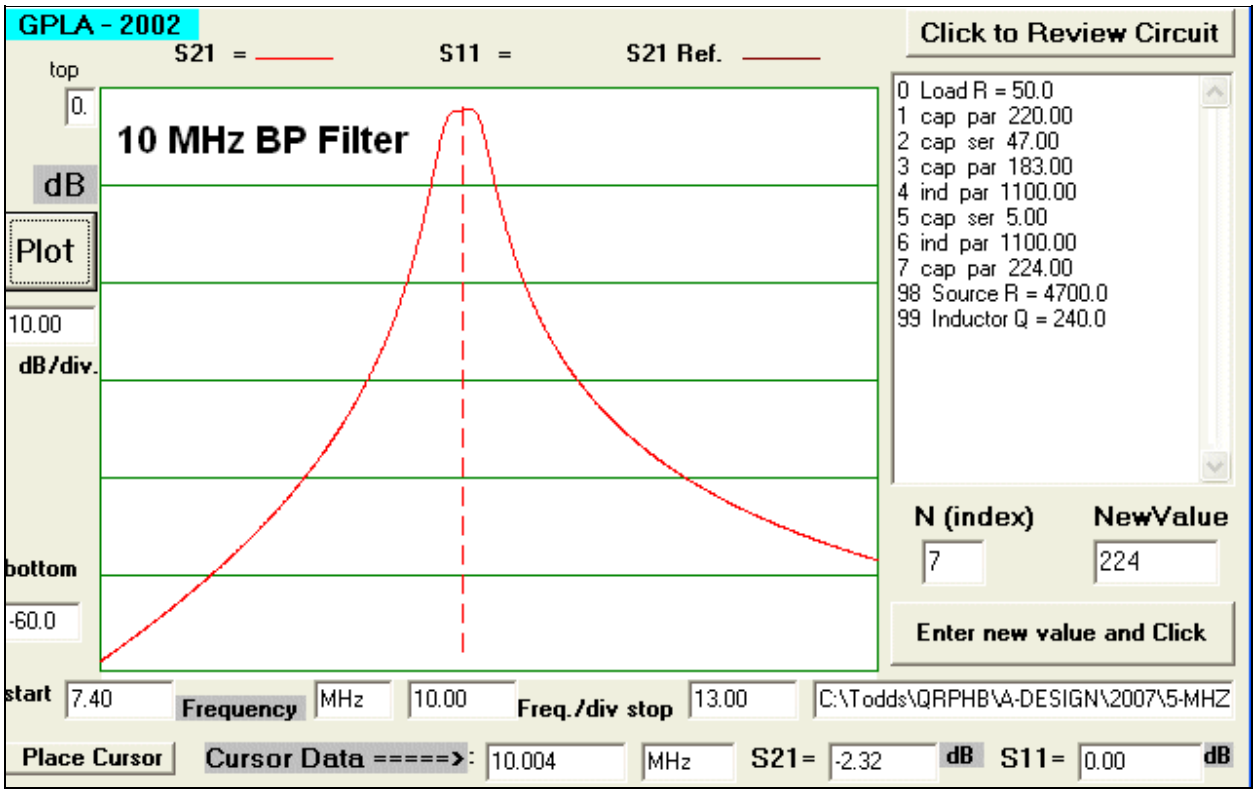
10 MHz Components

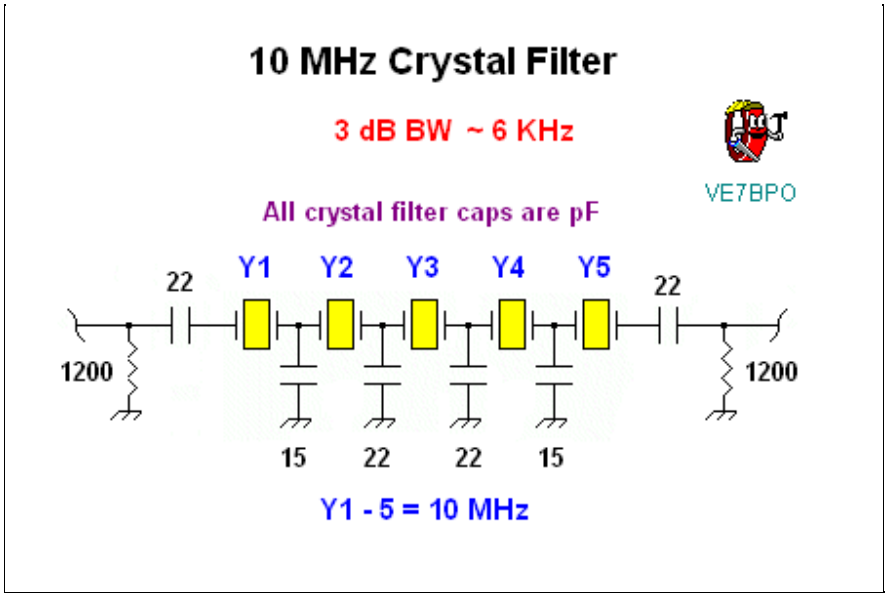
For those who wish to experimentally develop a 10 MHz version, I have included some potential band pass and crystal filter schematics you may wish to consider. In addition, my preliminary experiments indicated that another RF amplifier stage might be necessary for 10 MHz WWV. During development of these band pass and crystal filters, 10 MHz WWV reception was very poor and due to lack of time and the poor conditions, further experimentation was abandoned.
To the left is a 10 MHz band pass filter schematic with a -3dB bandwidth of around 100 KHz.



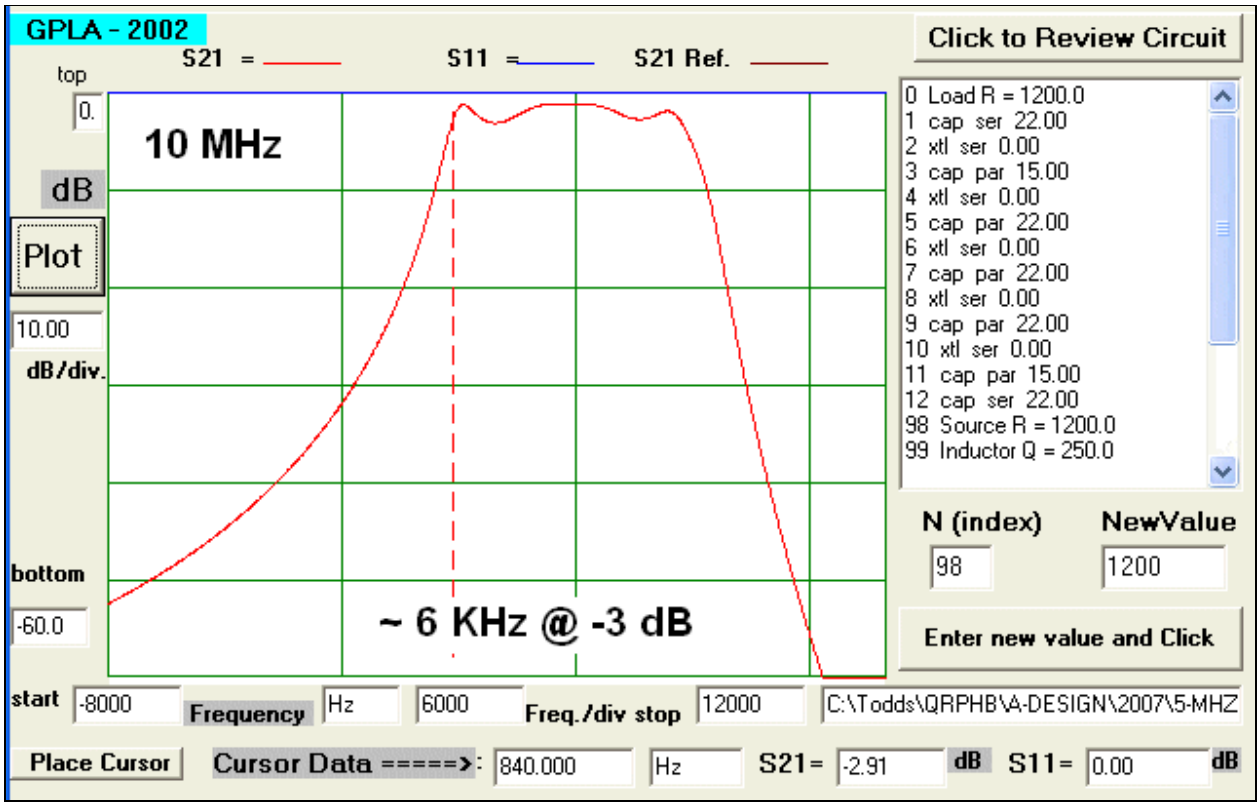


The GPLA simulation of the 10 MHz front end band pass filter. Tuning is very sharp with this filter.

To the left is a crystal filter design for 10 MHz. The realized bandwidth of these simple min-loss crystal filters is dependent on your crystal parameters.



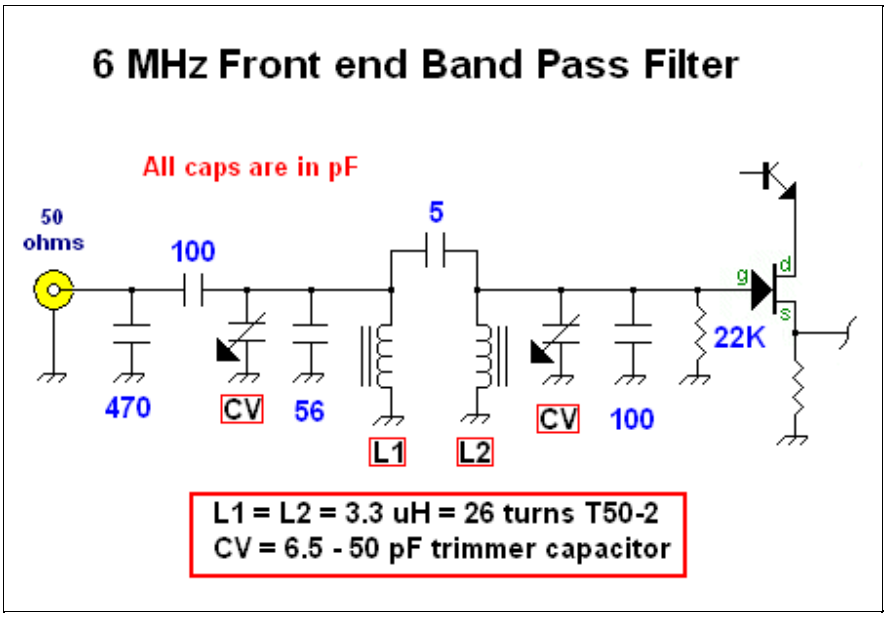
A photograph of the experimental 10 MHz crystal filter is shown to the right. Leads were kept long so that parts could be reused in future experiments.



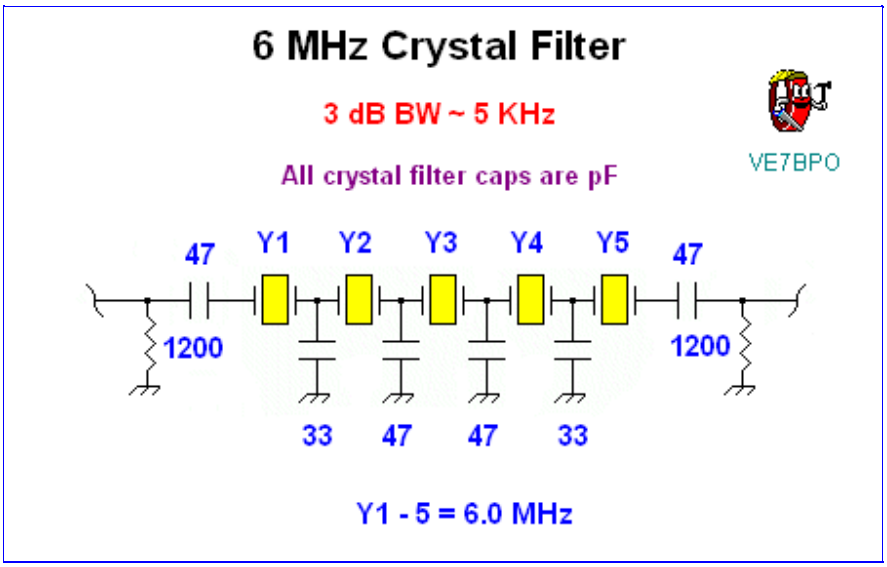
Above. The GPLA simulation of the 10 MHz crystal filter.

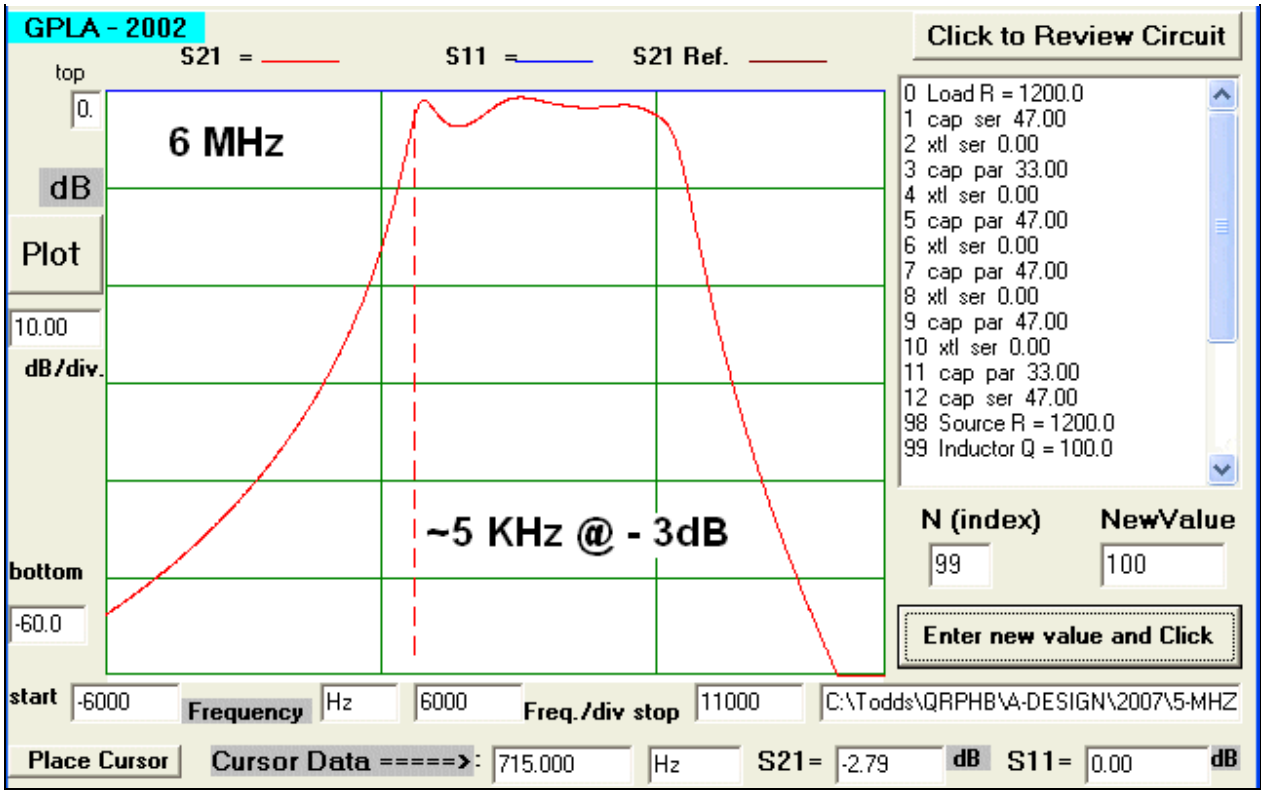
6 MHz Components

Since it is easy to find 6 MHz crystals and Radio Habana Cuba is on 6.0 MHz, band pass and crystal filters were designed for reference purposes. To the right is the front end, band pass filter schematic.



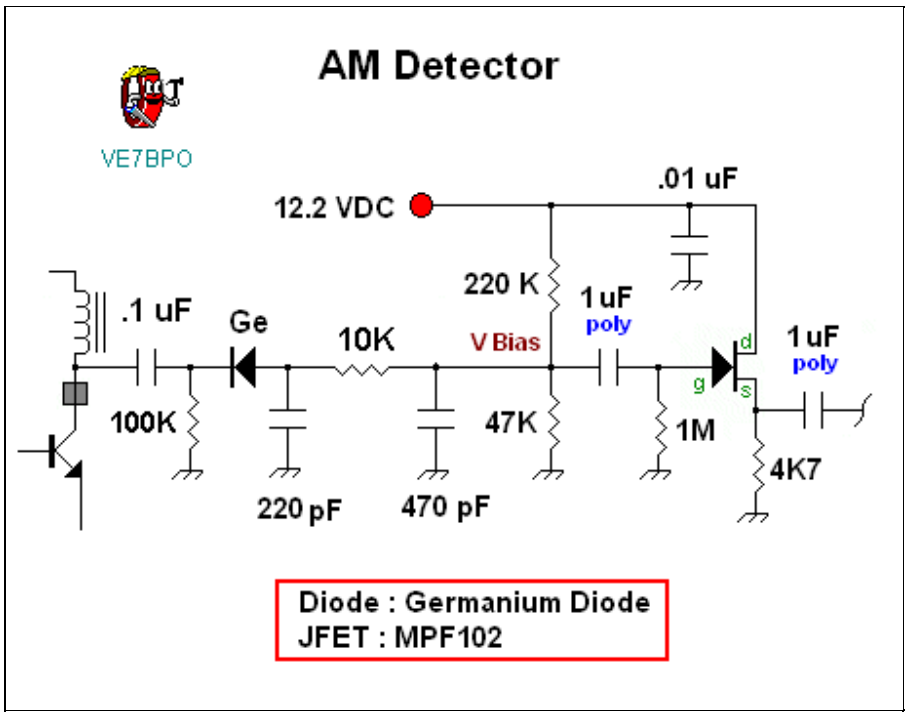
To the left is a potential crystal filter for 6 MHz AM reception. The bandwidth is ~ 5 KHz which is starting to get too narrow for some listeners, however, your crystal parameters will determine your filter's actual bandwidth.





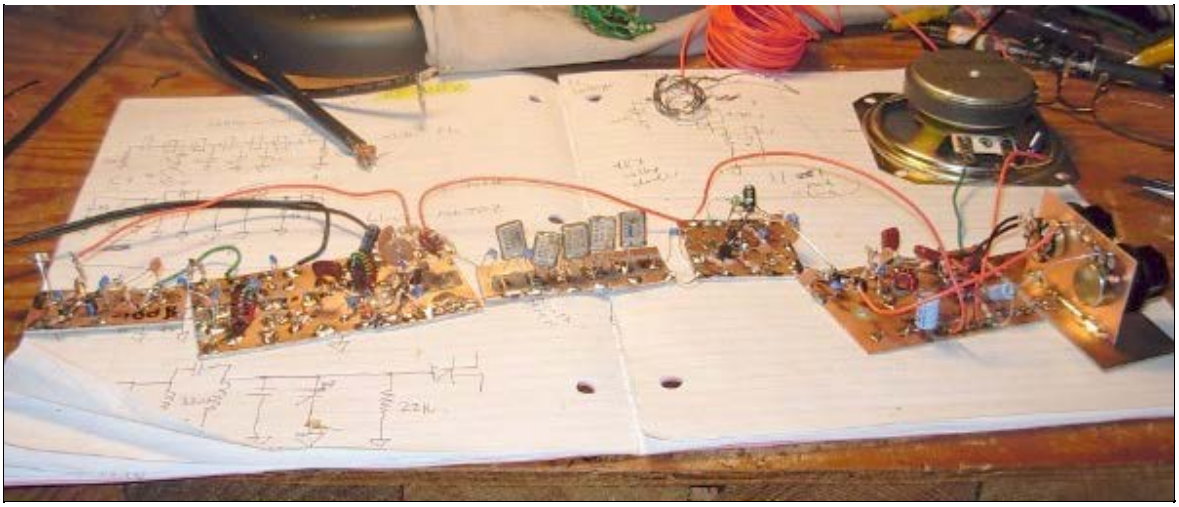
The GPLA simulation of the crystal filter using the parameters from the 6.0 MHz crystals I had in my parts collection.

VE7BPO Envelope Detector



To the left is the schematic of another detector experimentally developed for the 5 MHz WWV receiver. The germanium diode has incredible sensitivity when biased, although noise tends to increase with applied DC voltage. This basic design also worked well with a hot carrier diodes, although had less sensitivity. V bias can be changed by adjusting the 220K, 47K and 100 K resistor values. Note diode polarity.

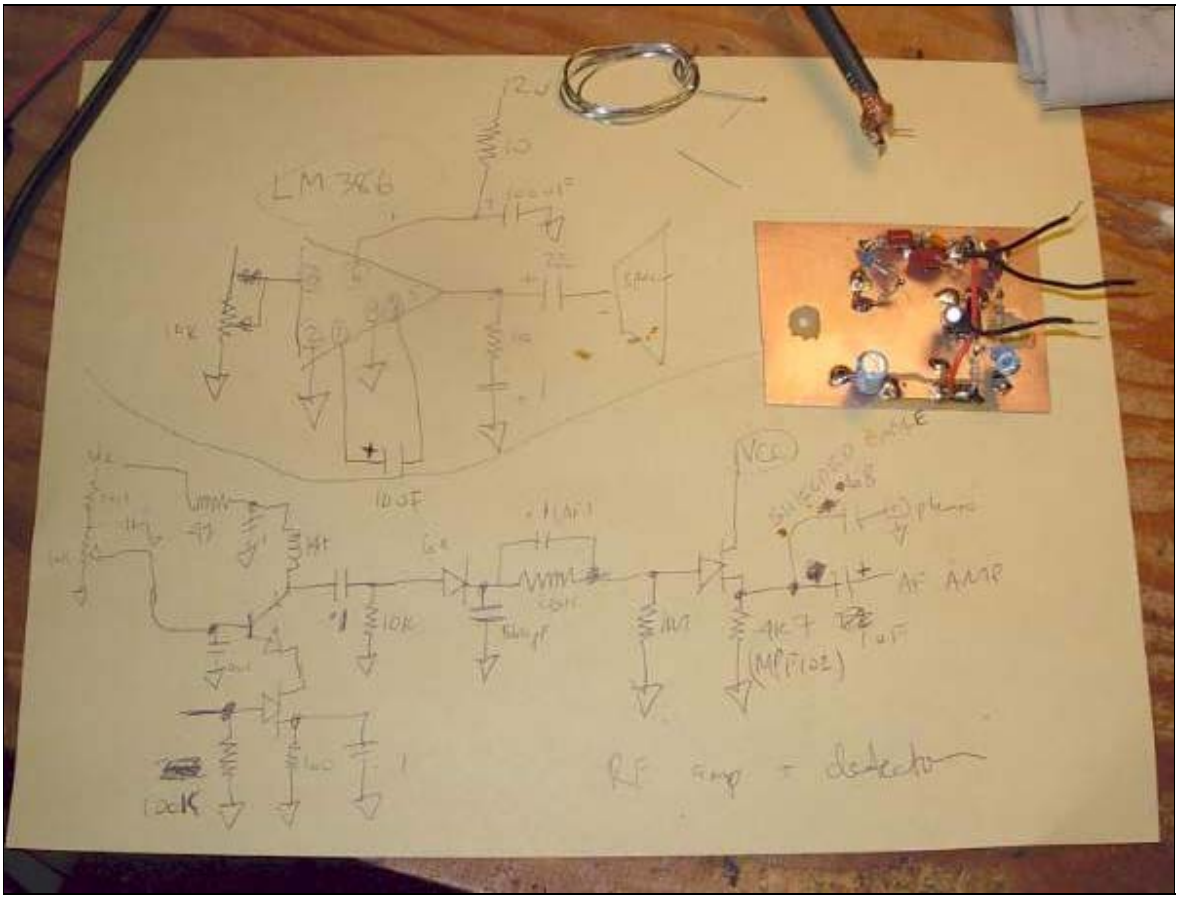
Miscellaneous Photos and Circuits



Above. A work bench version of the basic receiver bread board built in distinct stages from the antenna input to the speaker.

Above, A breadboard of the KK7B audio stage which is preferred to the LM386 stage used in the 5 MHz WWV receiver. The LM386 does take up less room and certainly increases the popcorn factor.

Above. The AF stage shown above in a chassis containing a speaker. This is my lab AF amp for testing receivers on my work bench.



A work bench photo of the first version of the W7ZOI detector built in October 2007 is shown above.

RF — Test and Measurement

My Links

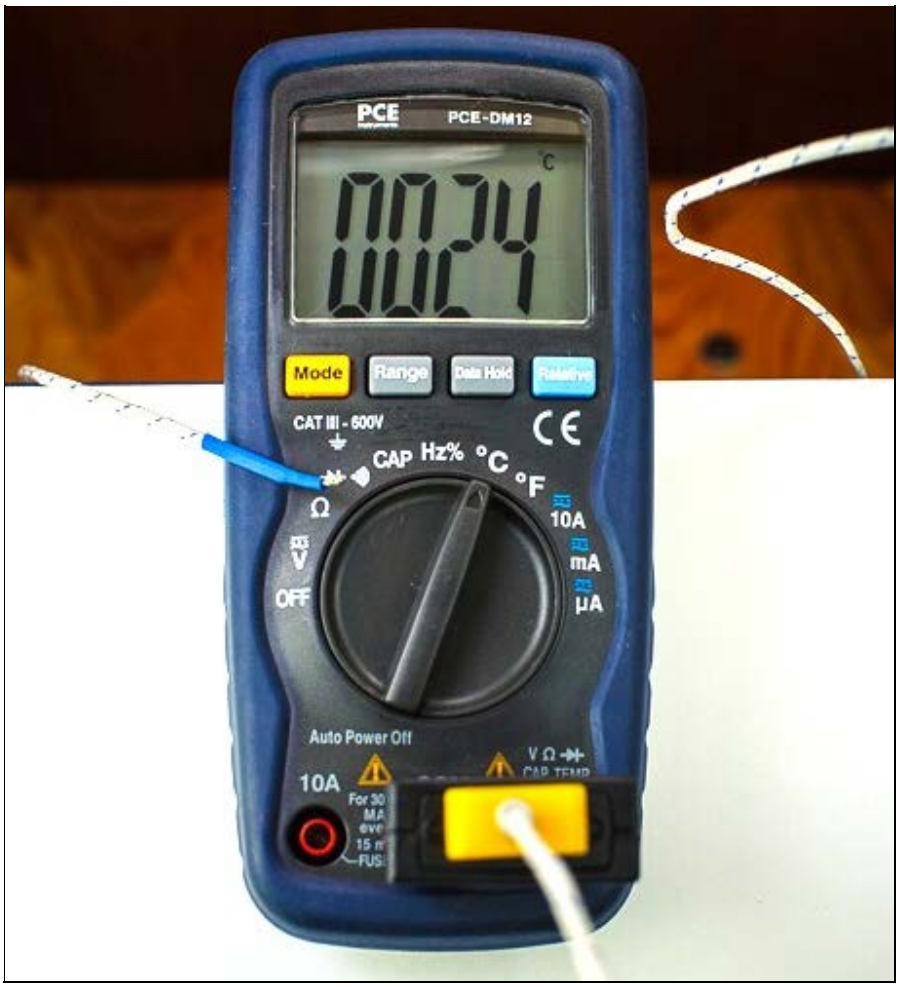
Greetings! I enter the QRPHB / SWL HomeBuilder site here and linked a few of my favorite tools, part suppliers and URLs I like and wish to remember. This page might feel intransigent.

In the following section called **Graphics**, I show the odd newer schematic or photo unpublished as permanent site content. A schematic might just represent my mistaken notion or whimsy: a sense of what's caught my fancy — not that my opinion is better than anyone elses.

Quick Navigator

Graphics

Russian giant sunflower — grew to over 3 meters tall



My gorgeous new PCE multimeter just in time for Season 16 experiments starting October 3. Click for another photo. Click for the PCE website — test equipment galore.

RF Transistor Data Spreadsheets

FET

BJT

Test Equipment

Test equipment

USA Rigol Dealer T Equipment

Online Tools and Publications

Standard resistor and cap values

Smith Chart information

Good inductive/capacitive reactance calculator other tools available

Resistive pad attenuator calculator

Ohm's Law 1 pager

VHF Communications website UKW-Berichte/VHF-Communications

The Signal Path EE video blog They review and show the DSA1030A SA + TG

DUBUS Technical magazine for 50 MHz and above

Favorite Experimenters

Esteemed QRP homebrew mentor - Wes Hayward's, W7ZOI web site

Iulian, YOUDAC-VA3IUL Comprehensive design information and amazing circuits web site

Markus Hansen, VE7CA A venerable Room 19 Radio School graduate. He has authored for QST. See his receiver featured in QST for March 2006

Blog of fellow VE7-land experimenter Steve, VE7SL. Look for a link to Steve's web site on his blog

Professor Ken Kuhn's web site. Perhaps start on his site map. A treasure trove of useful information

Ripples in the Ether An always interesting blog from our friend Jason who runs Etherkit

SV3ORA's Site Friend Konstantinos from Greece. Design info galore. Check out his microwave links.

FM Tuner Site Vintage FM Tuner Information

Paul, KE7HR Caver website

Matjaž, S53MV Microwave and RF. My favorite microwave designer. See this page for some inspiration

Paul Wade, W1GHZ authors my favorite QST column *Microwavelengths* His site

Jacques, VE2AZX web site Retired EE with great ideas

Michigan QRP Club Publish *The 5 Watter* Quarterly

Website of Michel, **F6FEO** from France. F6FEO project pages: click and click

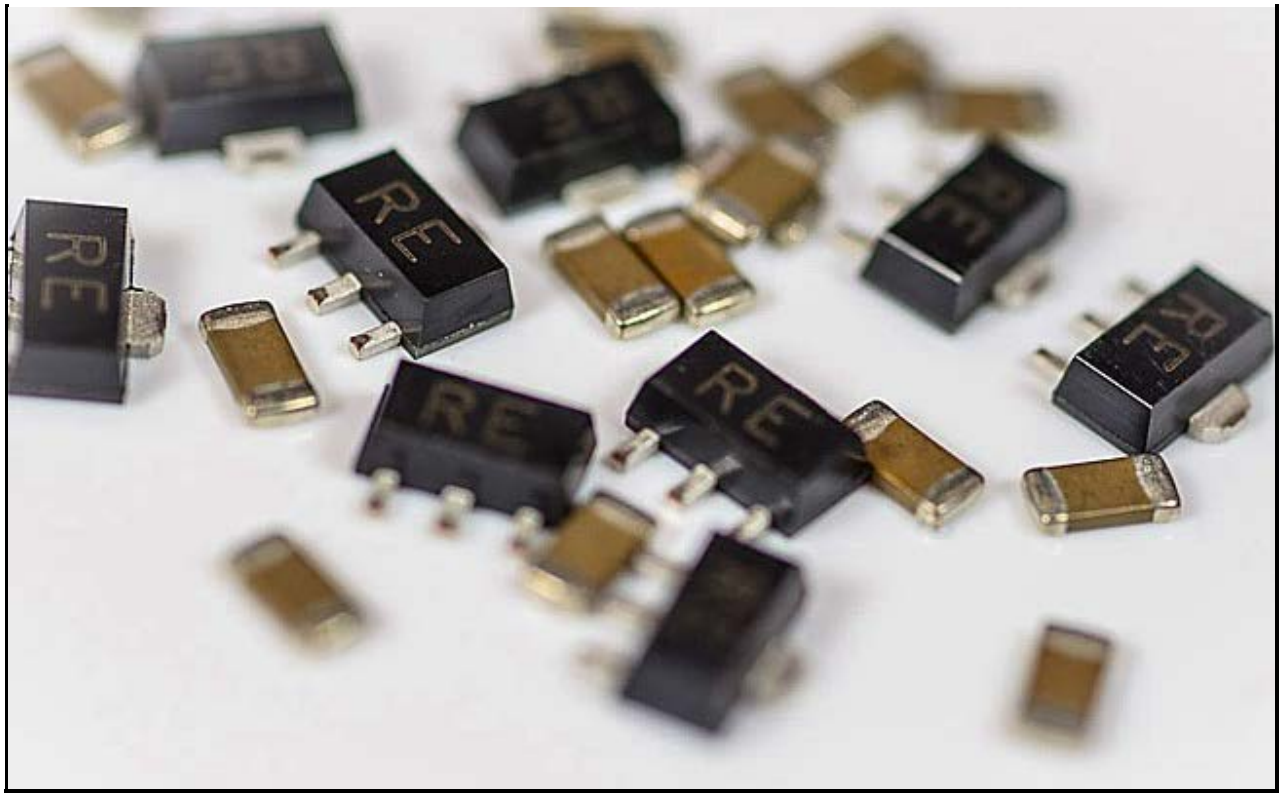
Roland, a retired EE who lives in France keeps this wonderful test equipment page

JF1OZL Mega site that first inspired me years ago. He once wrote me a letter - great guy

Brian, K6STI 88-108 MHz web site.

Leif, SM5BSZ Amazing work

KA7OEI Microwave Page web site


RE
RE
RE
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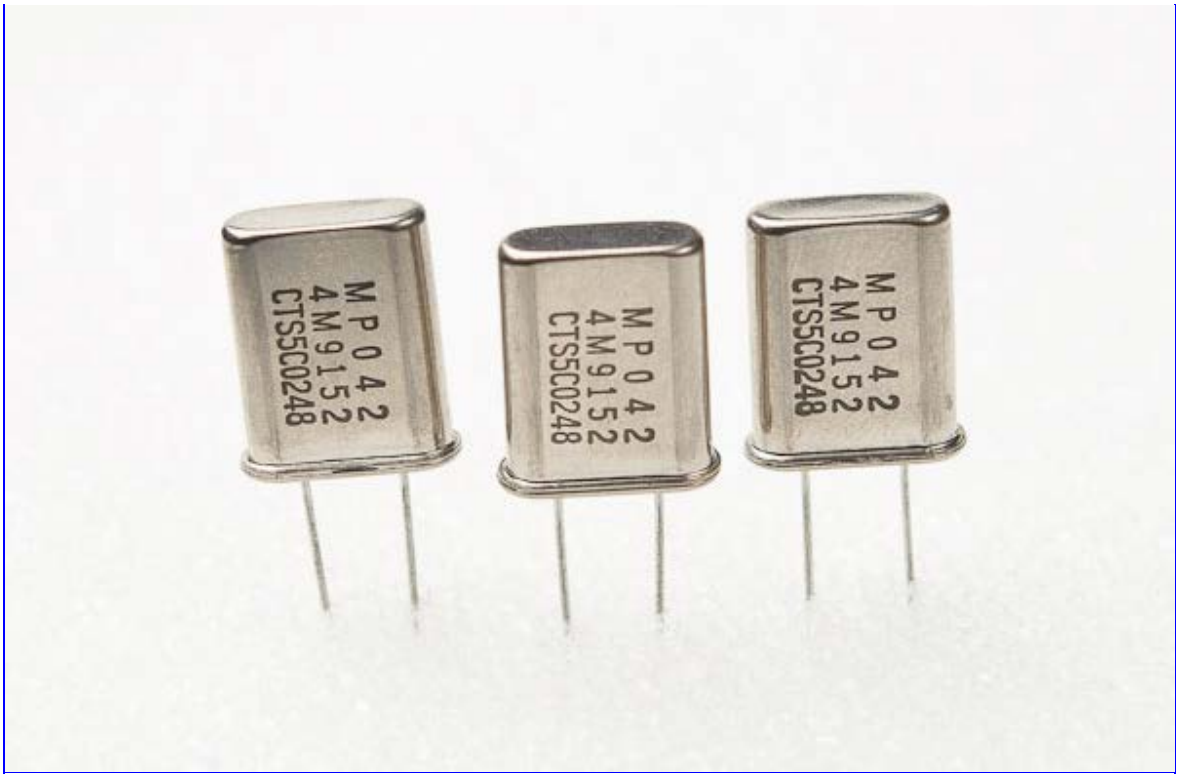
Memorial Museum of the Leningrad Siege, St. Petersburg, Russia
Государственный музей обороны и блокады Ленинграда

VE7BP0 photos 2006

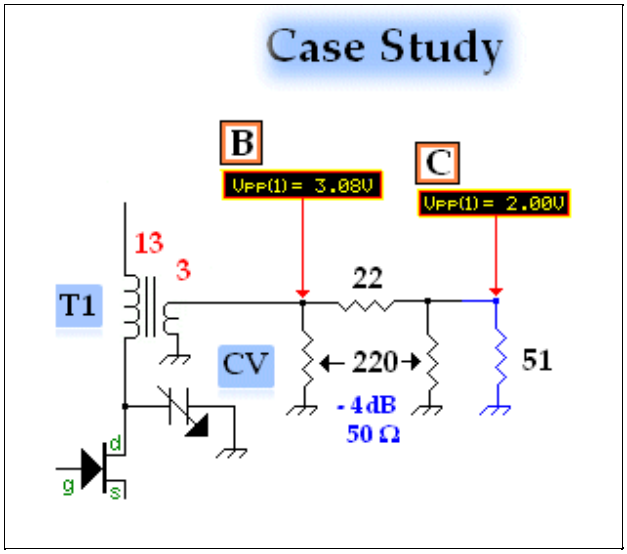

MP042
4M9152
CTS5C0248
MP042
4M9152
CTS5C0248
MP042
4M9152
CTS5C0248

RF Workbench Page 1 - Case Study Answers

This web page supplements RF Workbench Page 1

Answers:

1. Calculate the power in dBm at point B **= 13.75 dBm**
2. Calculate the power in dBm at point C **= 10 dBm**
3. What is the attenuation in dB of the 50 ohms attenuator pad? **= -3.75 dB**
4. What is the output power in mW of this stage? **= 10 mW**



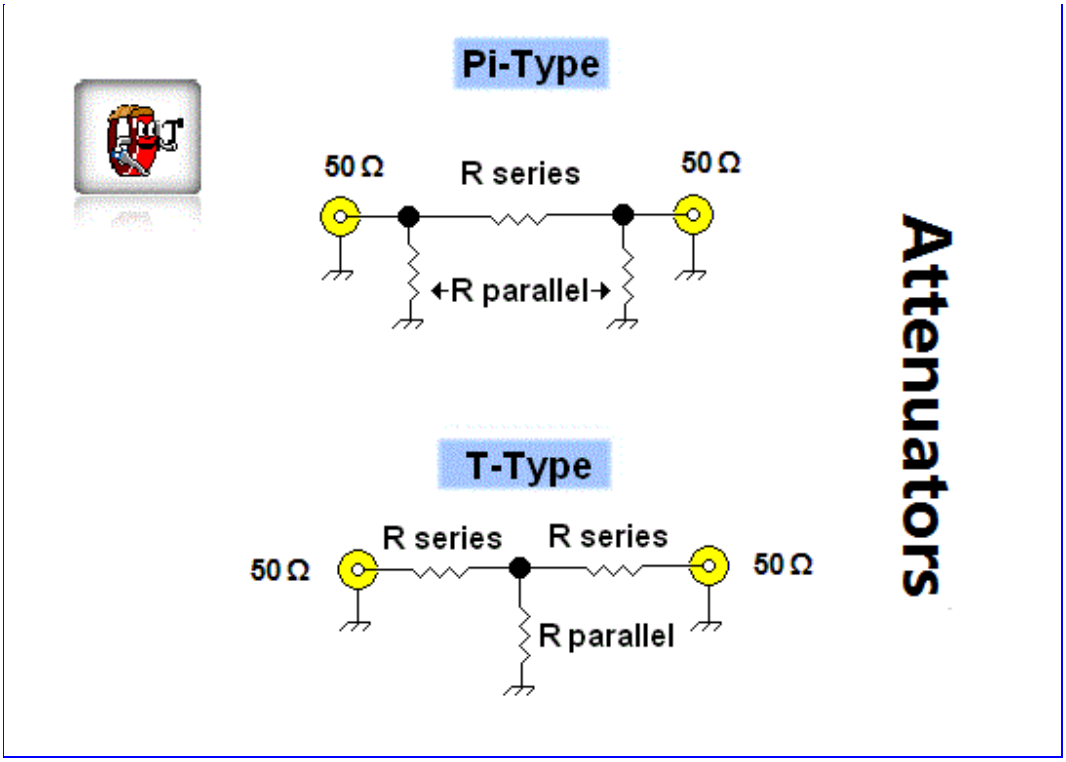
The design contained a "theoretical" - 4 dB, 50 Ω attenuator pad using standard value resistor substitutions

50 Ohms Attenuator Pads

This web page supplements RF Workbench Page 2

Table For Pi and T- type pads

Attenuation	Pi-type		T-Type	
dB	R parallel	R series	R series	R parallel
1	869.5	5.8	2.9	433.3
2	436.2	11.6	5.7	215.2
3	292.4	17.6	8.6	141.9
4	220.9	23.8	11.3	104.8
5	178.4	30.4	14	82.2
6	150.5	37.4	16.6	66.9
7	130.7	44.8	19.1	55.8
8	116.1	52.8	21.5	47.3
9	105	61.6	23.8	40.6
10	96.2	71.2	26	35.1
12	83.5	93.2	29.9	26.8
14	74.9	120.3	33.4	20.8
15	71.6	136.1	34.9	18.4
16	68.8	153.8	36.3	16.3
18	64.4	195.4	38.8	12.8
19	62.6	220	39.9	11.4
20	61.1	247.5	40.9	10.1
24	56.7	394.6	44	6.3
25	56	443.2	44.7	5.6



To build attenuator networks, we normally choose use nearest standard resistor values, or parallel/series 2 or more resistors from our resistor collection to approximate the tabled R. If you're building with 5% parts, consider measuring a few resistors with an ohmmeter to find the best match to the tabeled R.

Resistors in the 8 dB T-type or the 4, 14 and 19 dB Pi-type networks are close to 5% standard resistor values.

Expected Return Loss Table (50 ohm bridge)

This web page is a supplement to Transistor Input Impedance Experiments

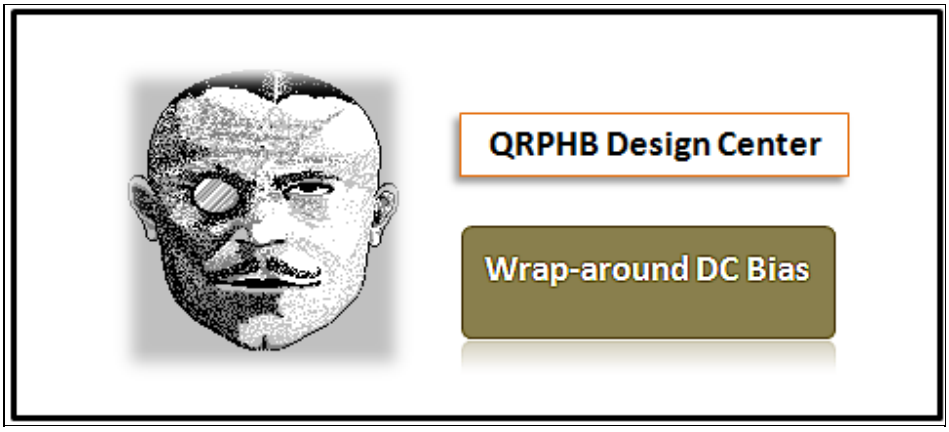
Expected Return Loss in a 50 ohm Bridge	
Resistor (Ω)	Return Loss (dB)
10.00	3.52
15.00	5.38
22.00	8.20
27.00	10.50
33.00	13.77
39.00	18.16
47.00	30.19
62.00	19.40
68.00	16.33
75.00	13.98
82.00	12.31
100.00	9.54
120.00	7.71
150.00	6.02
170.00	5.26
200.00	4.43
220.00	4.02
270.00	3.25
330.00	2.65
390.00	2.20
470.00	1.86
510.00	1.71

Miscellaneous RF Topics 2011 Supplement

Introduction

This web page supplements the Miscellaneous RF Topics 2011 web page.

1. Design Center - Wrap Around Bias



PNP Wrap-around Bias for an NPN RF Amp

There are entire books written on RF power amps. I don't pretend to be an expert and offer a simple approach to help move novice builders up the learning curve. Big thanks to Wes, W7ZOI for the theory and explanation for EMRFD Figure 2.16; the wrap-around PNP bias of an NPN transistor with grounded emitter. This bias circuit is fantastic and helps prevent you from burning up too many expensive medium power transistors such as the 2N3866 or 2N5109 during experiments and operation. From our emails, the following experiments were performed and information gleaned.

Since the NPN emitter terminal is grounded or near to ground on my web page examples, the current can really run high — Don't forget to use heat sinks and appropriately rated resistors on the NPN circuit when you really go for power.

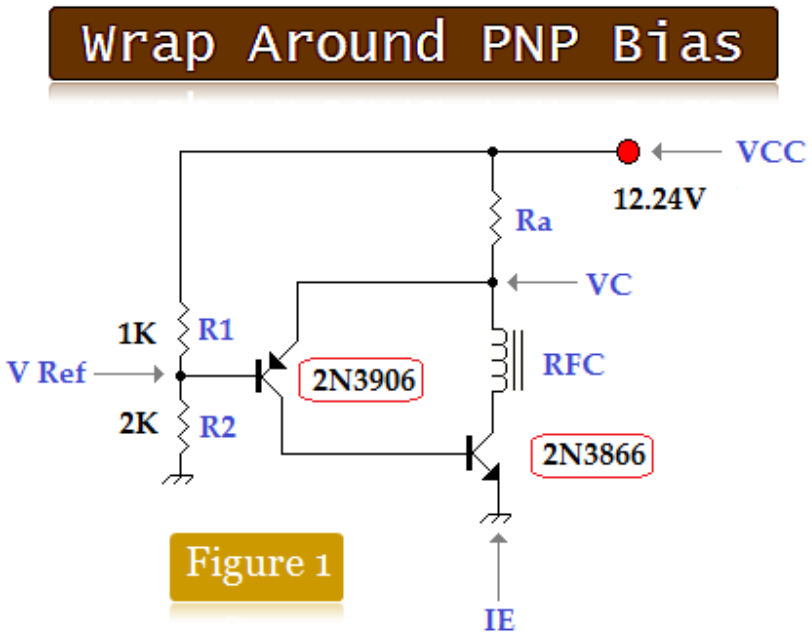


Figure 1

Desired IE	IE (mA)	VC (volts)	V ref (volts)	Ra (Ω)
19.0	19.83	8.62	8.15	180
42.0	43.51	8.31	8.15	82
105.0	108.50	8.22	8.15	33.3

Above — The schematic and data table from some biasing experiments. The normal decoupling parts were removed to isolate the biasing circuitry. All voltage and current values were measured. Another photo. A folded piece of paper is under the wire connecting the NPN emitter to the copper clad board. This is how I measure emitter current. The paper acts as an insulator and allows you to press hard with the ammeter probe to ensure good contact. After current measurement, the paper was discarded and the emitter soldered to the copper ground plane.

Biasing Procedure

Step 1: Choose a Target VC

Using collector resistor Ra, we want to drop about 2/3 of the power supply voltage across the NPN transistor. For a 12 volt supply or VCC, this is 8 volts VC. The formula is Target voltage = VCC x 0.66. Why choose 2/3 VCC instead of ½ VCC like we do for Class A audio amplifiers? In audio amps we just use a resistor as the load and the collector voltage can't go above the power supply unless we insert some reactances in the circuit. In simple AF voltage amps, the half VCC supply allows the greatest undistorted, peak-to-peak AC signal swing.

In RF power amps, a transformer or RFC lies on the collector terminal, so the instantaneous collector voltage (sum of the supply voltage plus the peak AC voltage) can go well above the supply voltage. If you have 8 volts on the collector, VC can easily rise to 16 volts or so on signal peaks. It's often prudent to choose a VC closer to VCC than the classic ½ VCC of audio amplifiers.

Hereafter, Target VC is referred to as **VRef**

Step 2: Choose R1 and R2 to bias the PNP @ VRef

Choose the PNP biasing transistors R1 and R2 so that the PNP bias = VRef. The formula follows:

$$VRef = \frac{VCC \times R2}{R1 + R2}$$

I wrote a Java Applet that does the math Applet A ; just plug in a VCC and play with the resistor values until you get your target VRef with the

resistors from your parts collection. Try to get a bias as close as possible to the targeted VRef, but don't stress out over it.

Step 3: Choose Ra to give (VRef + 0.6)

Ra = VCC – (VRef + 0.6) / desired current in amperes. Use or make up Ra with nearest standard resistor value(s) with an appropriate power rating. In a PNP, the emitter terminal is 0.6 VDC higher than the base, so need to we add that to VRef. The desired current is any reasonable current that you want to set up; for example 25 mA or 75 ma.

Example Calculation

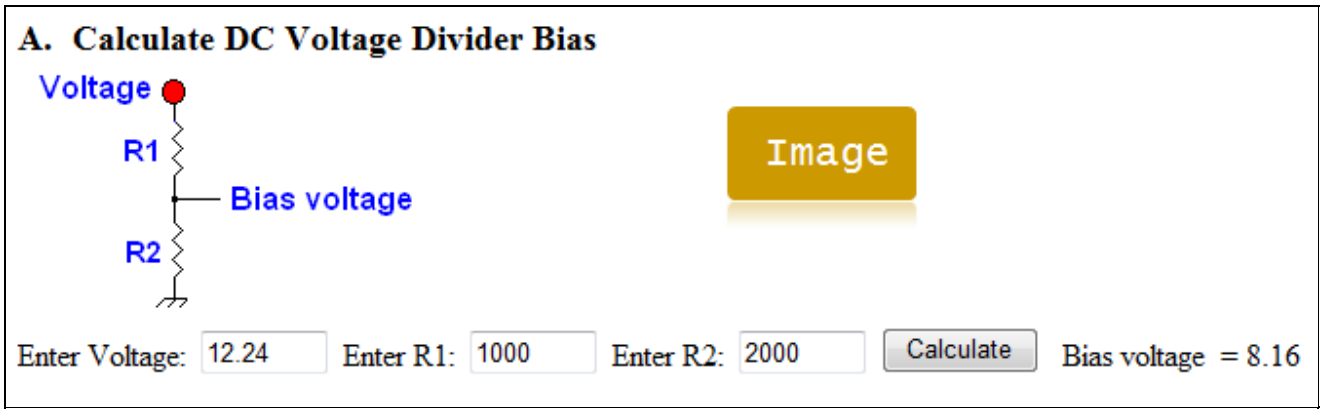
I wanted a emitter current of 105 mA from the Figure 1 circuit. VCC = 12.24. Let's run through the steps.

Step 1: Choose a Target VC

Target voltage = 12.24v x 0.66 = 8.1 VDC -> Now called VRef

Step 2: Choose R1 and R2 to bias the PNP at VRef

I just plugged some resistor values into Java Applet A. R1 is normally lower than R2.

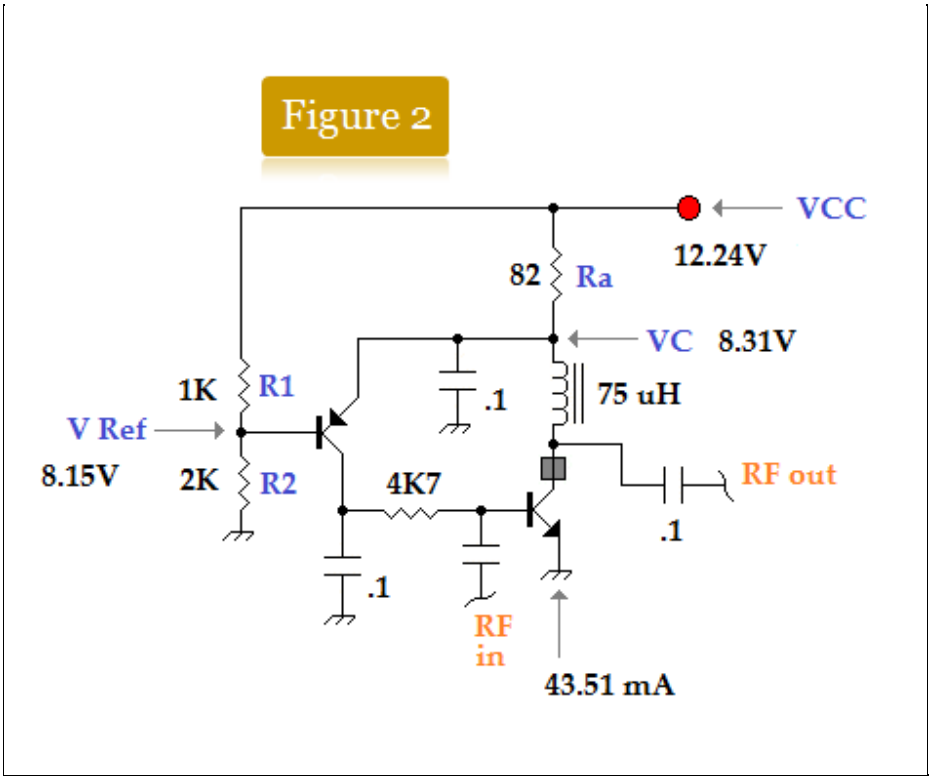


Above — A practical bias network. I have lots of 1 and 2K resistors. 8.16 v is close to my target VRef.

Step 3: Choose Ra to give (VRef + 0.6)

12.24v – (8.16v + 0.6v) / 0.105 A desired = 33.1 ohms. I put three 100 ohm ½ watt resistors in parallel for Ra. Click for lateral view photo showing the 3 resistors.

The exact measurements of this example and 2 other (19 and 42 mA) are tabled in Figure 1. I used an 82 ohm 1/2 watt resistor for Ra in the 42 mA version. The bench values closely approximate the calculated values. This circuit is quite instructive. In a real circuit (Figure 2) we add in the decoupling parts. Ra performs double duty as the decoupling resistor as well as the current regulator resistor.



Above — A working version of the desired 42 mA medium power amp from Figure 1. Refer to EMRFD Chapter 2 for further information.

How does the PNP wrap-around bias loop work?

Using the sample conditions from Figure 2: VC = 8.16v calculated, Desired IE = 42 mA and VCC = 12.24v
The main decoupling/current limiting resistor Ra is established at 12.24v – (8.16v + 0.6v) / 0.042 mA = 82.9 Ohms.

No matter what, the PNP VRef = 8.16 volts. All of the desired conditions are established and the current is flowing in the NPN. Let's change these conditions and see what happens.

First, instead of 42 mA, assume that the NPN current drops to something less. Hence, there will not be as much of a drop across Ra. This will cause VC to increase. This is the voltage on the PNP emitter. When this voltage increases, there is more voltage between the base and the emitter of the PNP, which will make it draw **more** current. That current comes out of the PNP collector runs to the base of the NPN. This causes the NPN to draw more current (beta times) which increases its collector current and causes the collector voltage to drop because of the I x R drop in Ra. It drops until it reaches the 8.16 volt level.

Let's now assume the opposite. The NPN draw too much current, more than the desired 42 mA. This means that there is too much I x R drop in Ra and VC goes **below** 8.16 volts. This tends to turn the PNP off. The PNP collector current drops, so the NPN base current also drops, causing the NPN collector current to drop. This causes VC to again increase to reach 8.16 volts. Is this cool or what?

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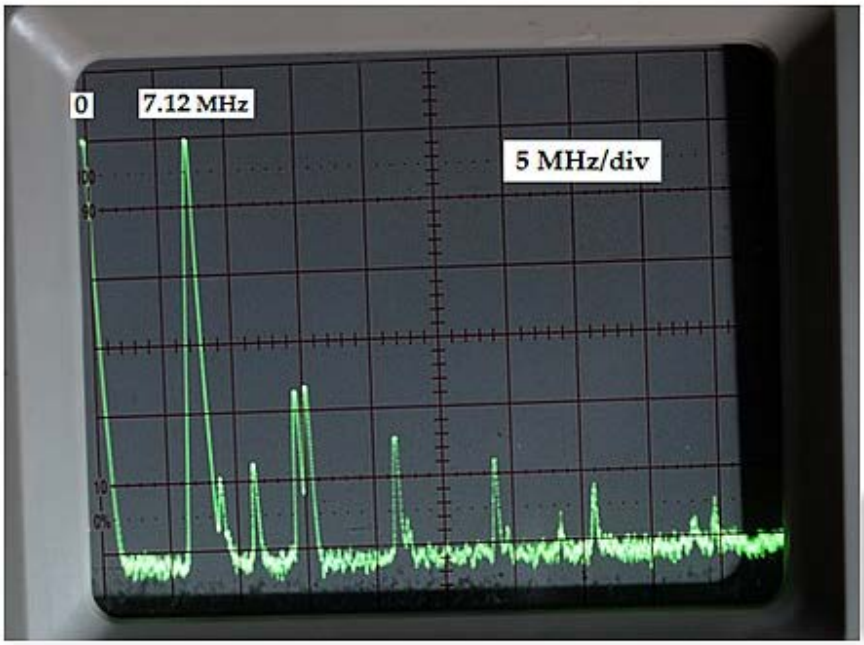
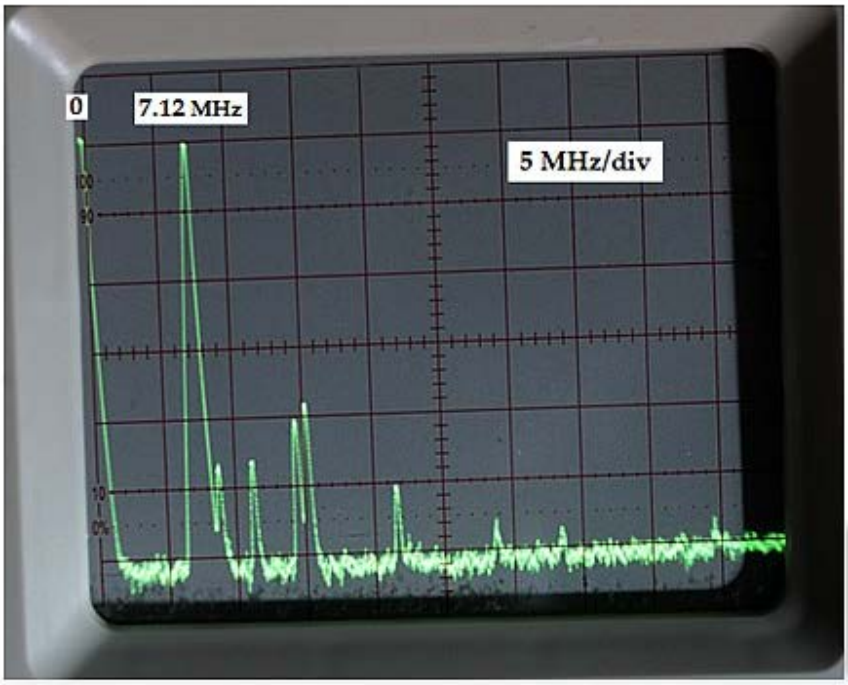
2. RC Low-pass Network

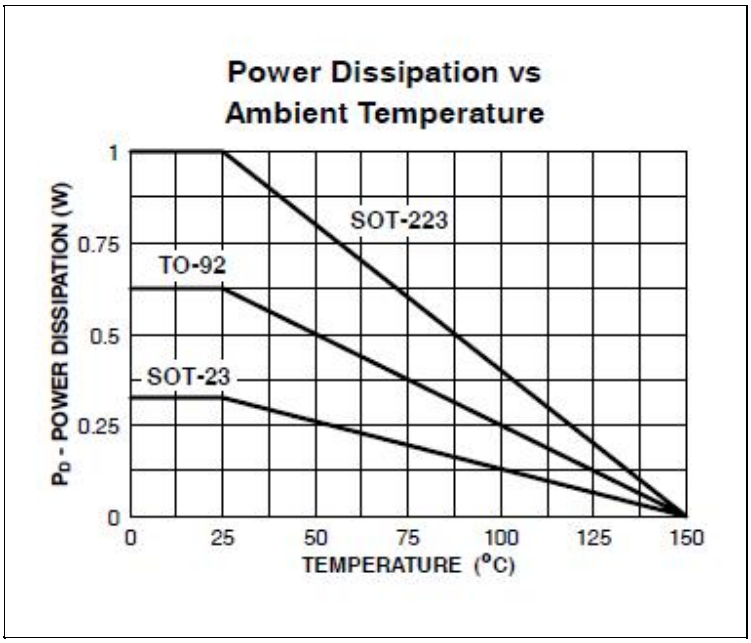


Above — Spectrum analysis without the RC low-pass filter.



Above — Spectrum analysis with the RC filter in-situ. Above 20 MHz, the spurs are ~5-6 dB down.

3. Transistor Power Dissipation in the 2N3904



Above — Power Dissipation versus Temperature for the 2N3904 in 3 packages.

Power dissipation in the collector resistor

For example, VC (the voltage across the transistor) = 10 volts and the emitter current = 20 mA

P = I x E = .020 amps x 10 volts = 200 mW

Power dissipation in the emitter resistor

P = I x E = .020 amps x the measured emitter voltage

or P = I x I x R

For example, emitter current = 20 mA with a 5 ohm emitter resistor; P = .020 amps x .020 amps x 5 ohms = 2 mW

We assume IE = IC in the above examples. At 200 mW collector dissipation, the 2N3904 temperature temperature climbs to about 65 C. A metal can transistor such as 2N2222 has more surface area and better heat dissipation. A heat sink also improves heat dissipation. Temperature dissipation charts are referenced to ambient temperature which may be dependent on the air circulation around the transistor. If you plan on running 200 mW or so collector dissipation with a 2N3904, glue on a heat sink. A pinch of the finger and thumb, or a infrared thermometer are the most common popcorn ways to test for the need for more heat sinking, or a part with better power dissipation.

Above — I learned this from a Russian builder. 100 heat sinks for a dollar. Typically, a small blob of epoxy with the transistor positioned to the lower edge of the penny works best. I used too much epoxy and didn't wait for it to completely dry before taking this photograph — the penny slid down.

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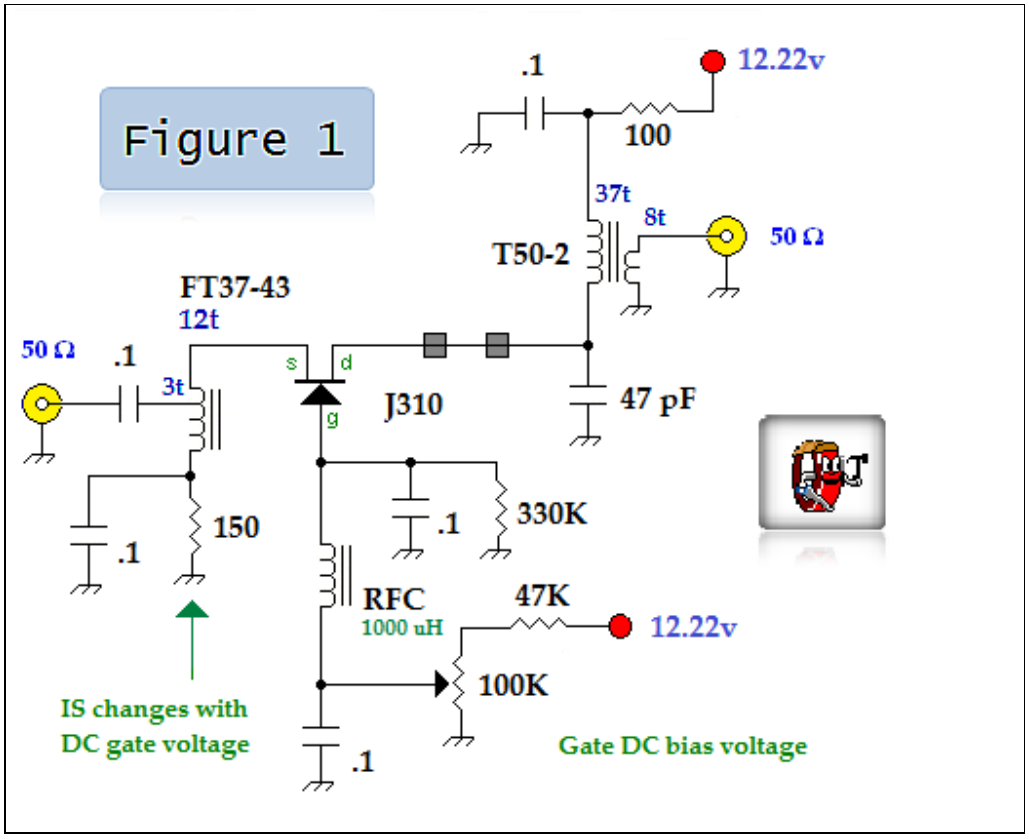
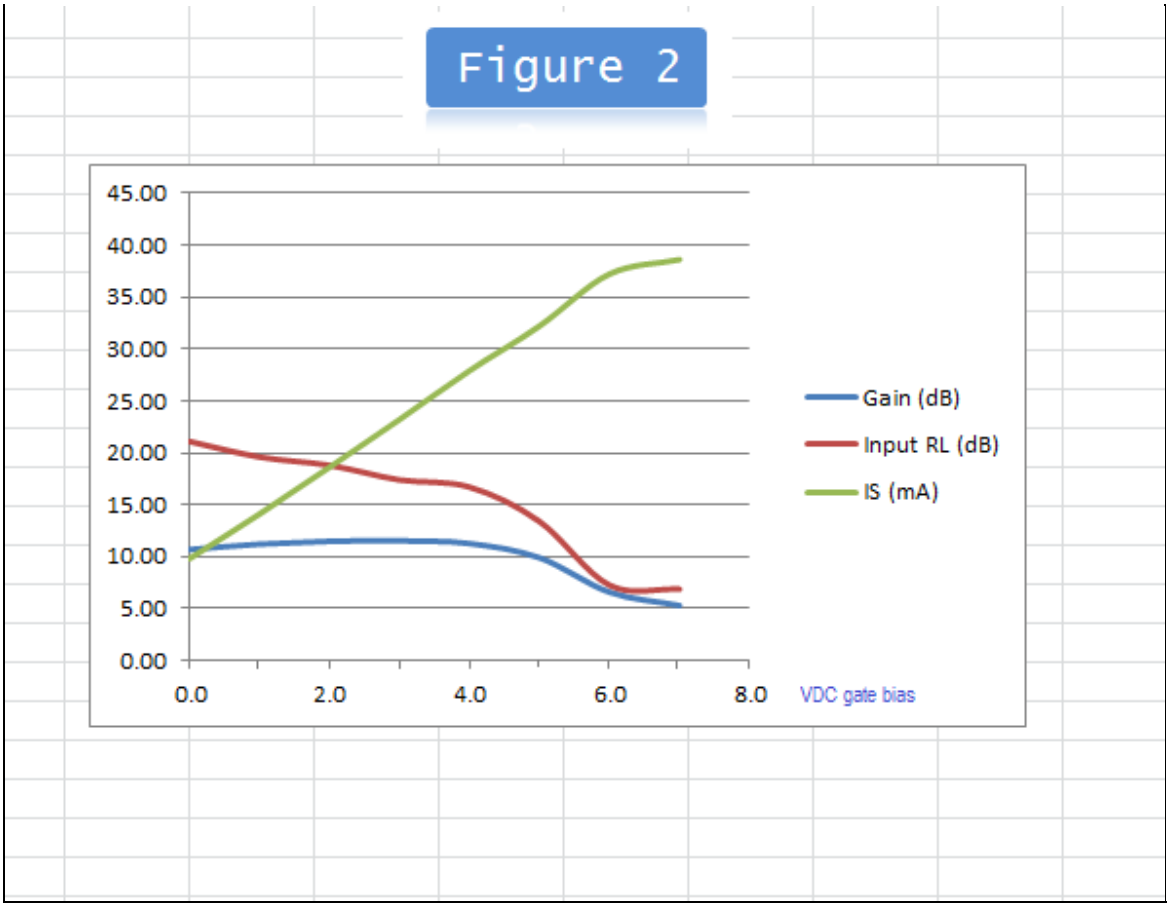
4. Some Factors Affecting Common-Gate Amplifier Input Impedance

A question arises: with all the devices and topologies available, why build a common-gate JFET RF amp in 2011 and on? For the J310 JFET at least: they're cheap, the input impedance is close to 50 ohms and wideband, the amp demonstrates good stability, signal handling capability and linear performance — all without running mega-current. Мне нравится. Many authors state the common-gate noise figure is inferior to that obtainable with the common-source topology, but for the popcorn builder, the common-gate noise figure is probably reasonable.



Above — A JFET gate biasing experiment. Here I applied 0-7 volts DC to the gate of a common-gate FET amp. The results proved interesting; see the graph below. The drain was tuned with a fixed 47 pF cap and manually "scrunch tuned" by pinching the T50-2 toroid with fingers and thumb. The input RFC serves to decouple the DC from the AC signal and the tap is used to improve the input impedance match.

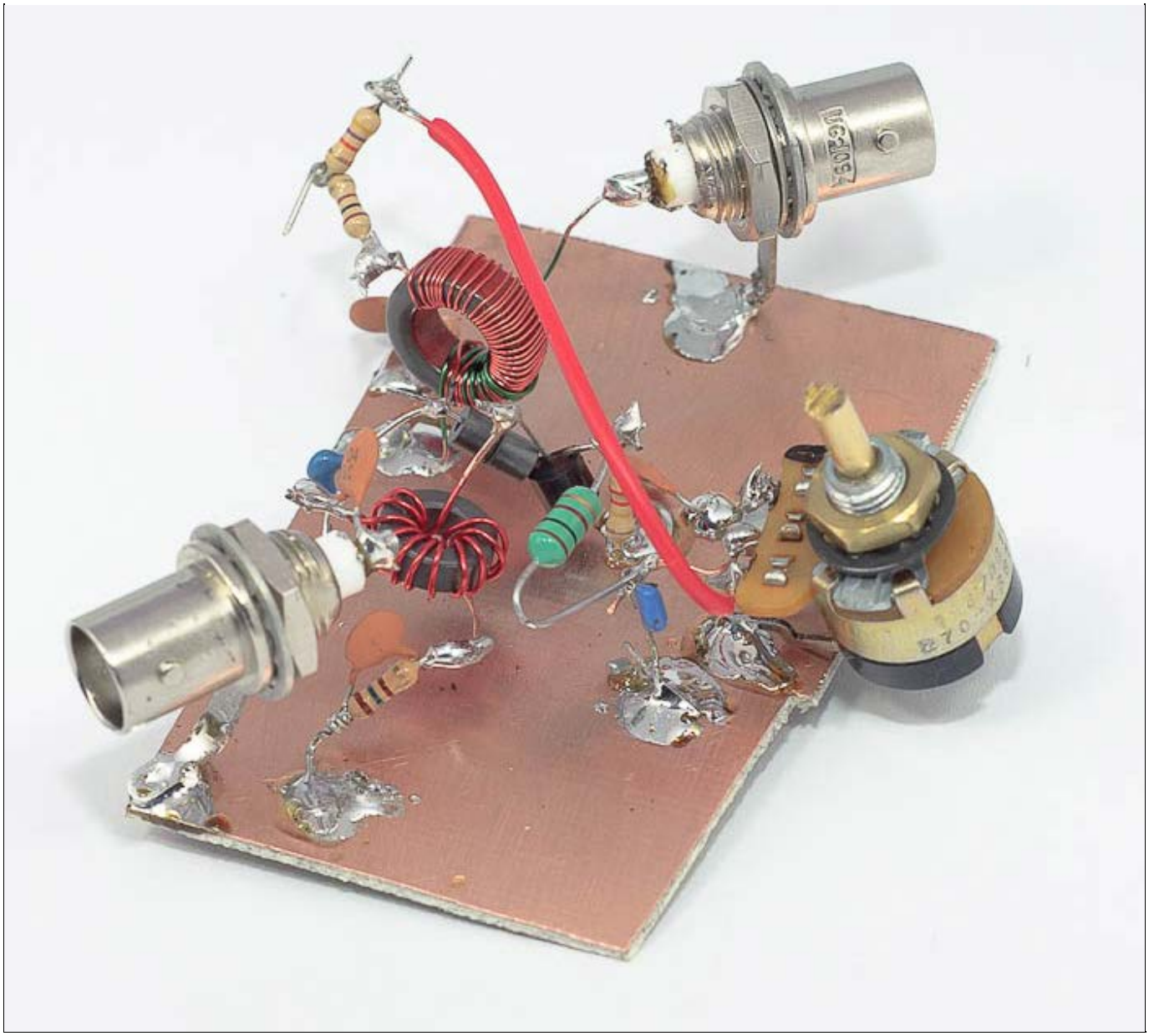


Above in Figure 2 — A graph of the DC gate voltage versus gain, input return loss and source current from Figure 1.

Increasing the DC gate voltage increases the source current as shown in the graph — ranging from 9.8 to 38.6 mA. From a DC gate voltage of 0 to about 4 volts, amplifier gain changes very little. Technically, the gain of this amplifier should increase along with drain current, however, the input resistance simultaneously goes down, creating greater voltage division loss at the input plus reduced return loss from impedance mismatch. This tends to cancel the increased gain as the plot shows.

The AC signal current injected into the FET comes out of the drain with power gain because that same current is now flowing into a higher impedance. The gain is dependent on the ratio of the impedance at the drain to that at the source. Click for a SPICE plot of DC gate voltage versus source current for 0 - 5 volts from Wes, W7ZOI.

From experiments, I learned that at least 6 things effect input return loss: the source resistor value, stacking or paralleling JFETS, the gate DC voltage(all which change current and/or transconductance), whether or not you bypass the source resistor, negative feedback and also the tap on the input RFC. Like other amps, the impedance of the device before and after the JFET amp can also change the input match, but those are external and normal considerations.



Above — Breadboard of the DC gate bias experiments of Figure1. Properly biased JFETs act like a current source where drain current is controlled by the gate-to-source DC voltage. The input impedance of the common gate amp is the reciprocal of the transconductance. Ken Kuhn posted some great JFET design notes in the yellow JFET section on his web site. EMRFD provides another essential reference for RF JFET design.

Noise Figure and the Common Gate Amplifier

Noise figure is important in low-level RF amplifiers such as a preamp or post-mixer amp because noise generated by these amplifiers contributes to the overall noise figure of a receiver. The input impedance an amp like the common-gate is considered an *active* impedance. Often, with active impedances, obtaining the maximum power transfer and lowest noise figure don't occur simultaneously. That is, the optimum input impedance match for gain is not the optimum input match for low noise performance. With pure resistances, the best noise figure possible is 3 dB; however, when using an active impedance, it's possible to achieve a noise figure less than 3 dB by applying negative feedback or other techniques to manipulate the JFET active input impedance and lower its voltage noise.

A stellar example by Bill, W7AAZI employs DC gate bias, noiseless negative feedback and paralleled JFETs to achieve a noise figure in the 1.2-1.5 dB range. This mind-boggling circuit is described in EMRFD as Figure 6.94. I wanted to explore this topic: can amateur builders design *common-gate* amplifiers for a lower noise figure? The answer is perhaps, however, unfortunately, there is no cookie-cutter approach.

Low noise amplifier design imposes an advanced topic and many builders lack the math skills and/or test equipment to design and generate results. Each design must be approached on a case-by-case basis, although similar fundamentals apply to all cases — using active impedance manipulation techniques and paralleling up JFETs while paying attention to all the resistances in a circuit — even loss resistance in the inductors. I learned from Professor Ken Kuhn that when amplifiers are paralleled the output signal power adds linearly, whereas the internal noise adds statistically (square root of the sum of the squares) and thus noise figure improves. In short, paralleling JFETs reduces voltage noise,

but only to a point, as input capacitance also increases which can degrade high frequency performance and noise figure. Ideally, the JFETs paralleled should be matched for IDSS and VP.

Further, all devices have a voltage noise specifications (in volts per root Hz) and a current noise specifications (in amperes per root Hz). The source impedance that produces the lowest noise is the ratio of the two — derived from Ohm's law. The input impedance of the device may be higher or lower; thus the optimum source impedance for low noise is not necessarily the optimum source impedance for maximum power transfer. It's unlikely that these two impedances differ widely, however, only careful bench measurements or closely copying proven designs will ensure the desired results.

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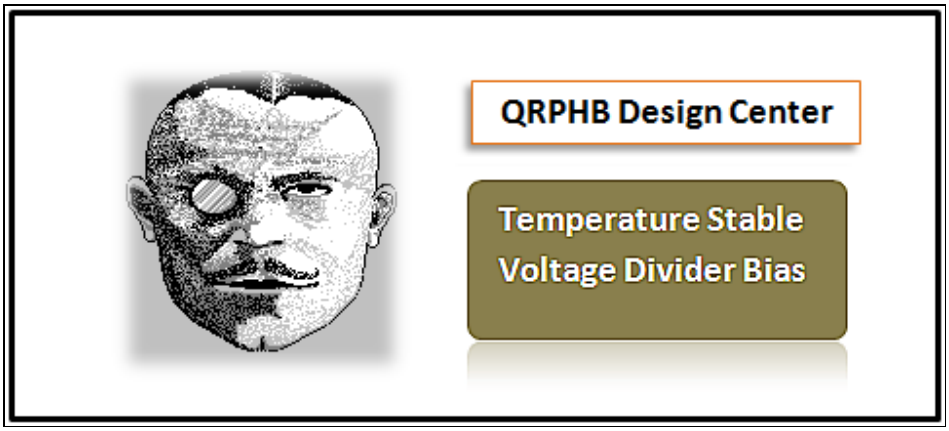
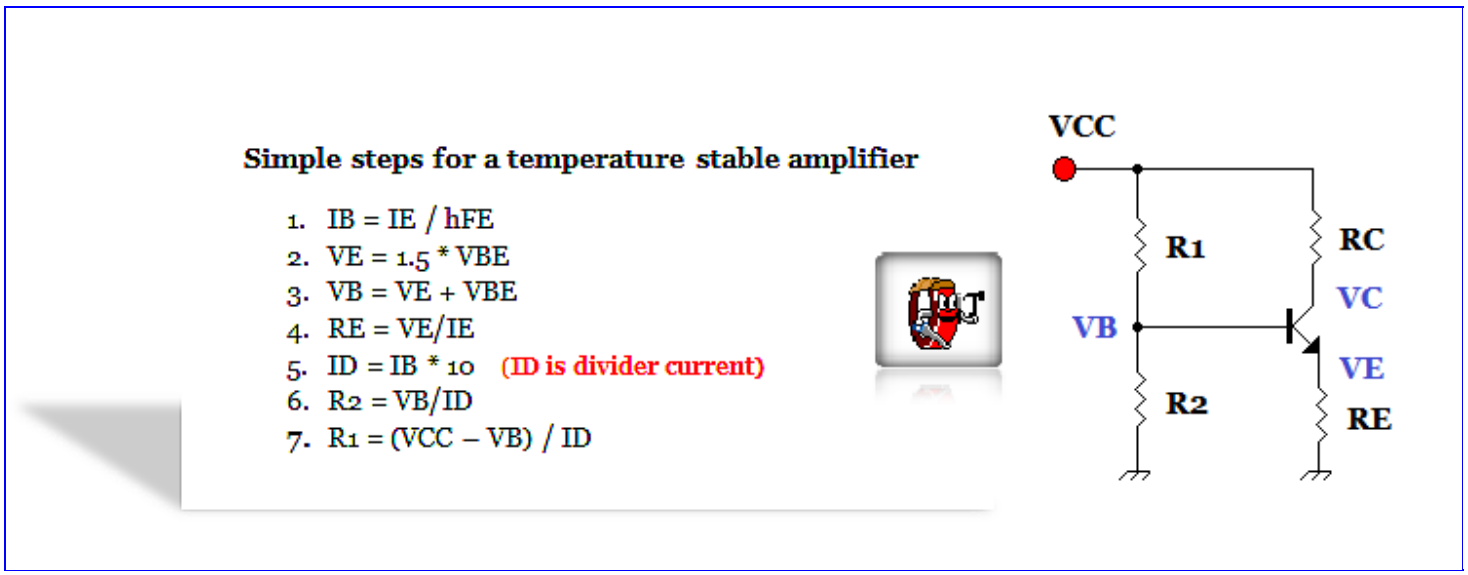
5. Design Center - Temperature Stable Voltage Divider Bias





Above — A simple way to design a temperature stable BJT voltage divider bias. Assumes that IC = IE. You specify IE.

I sought to develop a simple way for builders to design a temperature stable BJT amp using voltage divider bias. It wasn't so easy. Here is the full math from Ken Kuhn. As shown, it gets complicated and over-the-head of the target readers of this web site.

My simple approach makes assumptions, and certainly an astute builder could improve my algorithm - I developed other formulae, however settled on this one because it well fits the scope of the QRP/SWL HB web site. Most engineers design their bias networks in SPICE and I recommend this.

To choose a Beta or hFE, you can measure the Beta or use the transistor datasheet Beta near to the emiiter current you intend to apply. Many builders debate this and even go as far to state that spec sheet Beta values are useless. This may be true, but you *have* to decide on some approach. Ultimately, you will build the amplifier and can tweak 1 or more of the bias resistor values to set your bias voltage (emitter current). Click for a snippet from a 2N3904 datasheet showing hFE at a range of currents.

Biasing BJTs is a great example of building to suit your design requirements. The average reader likely operates their radios at room temperature where a minimum temperature stability factor is needed. On the other hand, you may operate field portable, or use a crammed chassis where transistors heat threatens stability. As always, your design choices must fit your needs, parts and abilities.
The above 7-step method forms a simple **QRPHB Design Center** method to achieve a reasonable temperature stability factor at normal

operating temperatures. I considered 2 factors: VBE and Beta (hFE) changes.

VBE

We normally attribute the voltage difference from base to emitter (VBE) as 0.7 volts for silicon transistors. VBE changes inversely with temperature and at AF to HF, we mitigate VBE creep by applying an emitter resistor to increase the emitter voltage VE. The emitter resistor provides a feedback loop so as collector current starts to increase due to a rise in temperature, the voltage drop across RE also starts to increase. This voltage tends to reverse bias the base-emitter junction. The net effect is decreased collector current.

How do I choose an emitter resistor value for VBE bias stability asked the little grasshopper (novice)?

The VE value should typically be about 10 to 25% of VCC. VE recommendations really vary from author to author — some say VE should be at least 5X VBE while other authors suggest 2-4 volts as the minimum acceptable VE value for bias stability in typical AF-HF amplifiers. I use 1.5 * VBE (0.7) = a VE of 1.05 volts, but it's up to you. Since you know IE for your transistor, calculate the resistor RE to provide the target VE using ohm's law: RE = VE/IE.

At VHF on up, emitters are typically directly connected to ground since that path must have as near zero inductance as possible. In these amps, the bias is derived by a feedback circuit that controls the base current to stabilize the collector current at a specific level. Inductance in the emitter lead causes significant loss of gain at higher frequencies and also upsets the input impedance to the base. Similar effects occur in JFETs and even vacuum tubes. 10s of nH of inductance at 144 MHz represents a reactance of 10s of ohms. At UHF on up this becomes very serious — that's why construction methods are so critical for these builders.

In conclusion; to influence VBE stability against temperature changes and to some extent leakage current, at AF- HF, apply series feedback to a raise emitter voltage (VE). Accounting for variations arising from 5% tolerance parts, I chose a minimal VE of 1.05 volts (equation 2 of 7 in my "Simple Steps..." algorithm).

A tradeoff challenges us — increasing the emitter resistance, increases AC degeneration, lowers AC signal gain and raises the transistor input impedance. Bypassing all or part of the emitter resistance offsets these problems.

Beta

Beta variation arising from temperature changes must be mitigated. I briefly discuss the factors that change Beta in the bullets under the #5 Bipolar Junction Transistor Beta Tester on this page.

To design "beta independent amplifiers", the builder alters the resistance ratio RB/RE. Decreasing this ratio improves the Beta stability while decreasing the amplifier current gain. The ratio can be confusing, so the Design Center equations 5, 6 and 7 just make the current in the divider 10X the base current. It's simple — you're done. Feel free to choose whatever ratio you think is best. Good values for the RB/RE ratio lie in the 12 - 20 range.

Design Example

[hFE = 100, IC = 10 mA (0.01A), VCC = 11.22v, VBE = 0.7]

*I used a 100 ohm low-pass resistor; so VCC = 12.22v - (0.01 A * 100 ohms) = 11.22v*

1. IB = IE / hFE = .0001 A
2. VE = 1.5 * VBE = 1.05v
3. VB = VE + VBE = 1.75v
4. RE = VE/IE = 105 ohms **(Use 82 – 100)**
5. ID = IB * 10 = 0.001 A
6. R2 = VB/ID = 1750 ohms **(use IK8)**
7. R1 = (VCC – VB) / ID = 9470 ohms **(Use 10K)**

Above — A design example employing my "Simple steps for a temp stable amp" bias algorithm. My VCC = 12.22 volts, but, as always I decoupled and bypassed the VCC — in this case, I applied a 100 ohm /0.1 uF network. At 10 mA emitter current, the voltage drop across the 100 Ω resistor results in a VCC of 11.22 volts.

Big thanks to Wes, W7ZOI and Professor Ken Kuhn for helping me with my math problems and to better understand biasing concepts.

A friend made a Design Center spread sheet for MS Excel. In this application you enter the desired IC rather than IE.

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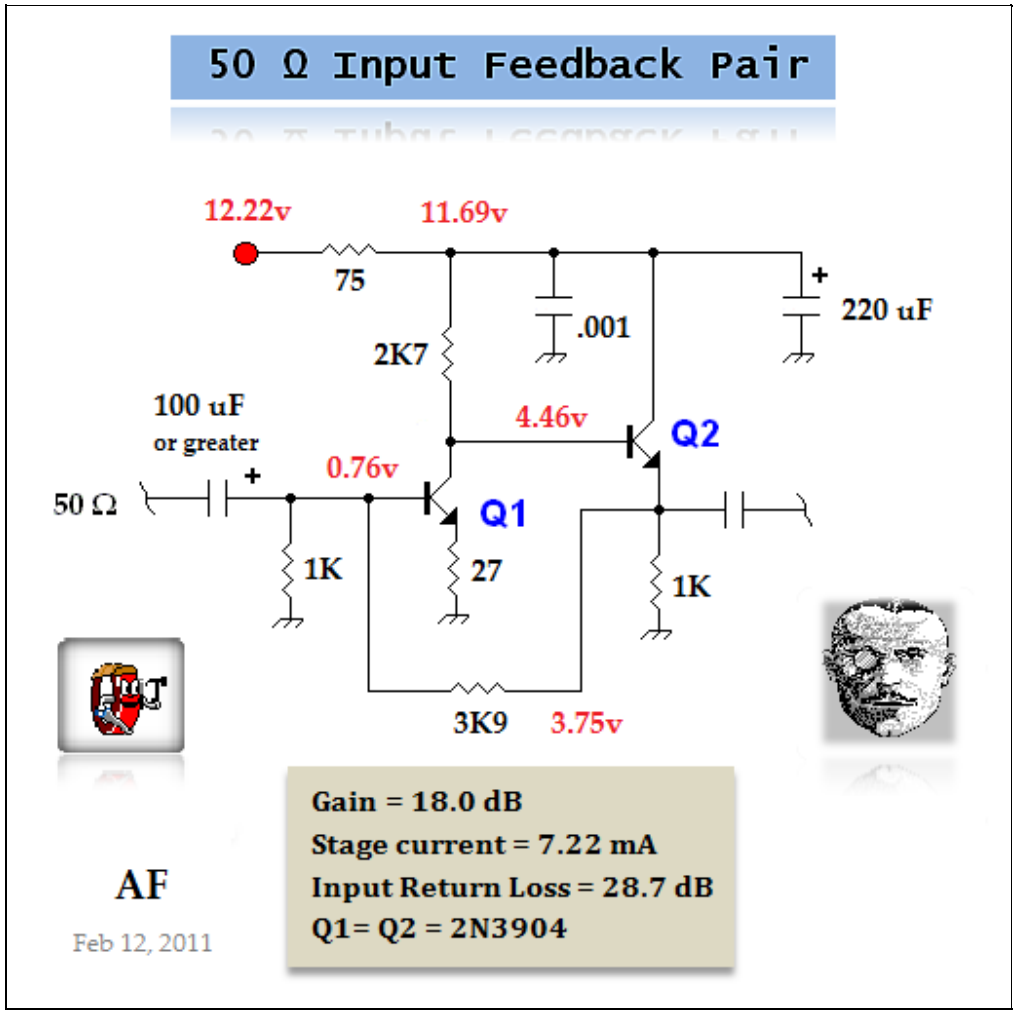
50 Ohm Common Emitter Audio Preamp Experiments

Introduction

This page supplements the Pop DC2 Receiver Experiments from this web page.



Above — A 50 ohm audio preamp to follow a diode ring product detector.

Many seek an alternative to the common base AF preamp that typically follows a diode ring product detector. Winter experiments yielded some possible ideas. The simplest is shown above.

Using all different parts, I built and measured another version to ensure reproducibility of the input return loss while just using standard 5% resistor values. The second amp proved similar: 18.4 dB gain, 7.27 mA stage current (3 .8 mA is consumed by the emitter follower), a return loss of 27.6 dB and a nice sine wave output. A 2K2 resistor AC coupled to the emitter follower provided a load. Note in the photo, I used 1% resistors

in some slots, but the values were standard 5% and not unusual values like 15.4 K etc.. The 3K9 feedback R is the most critical value — it's a 5% part.

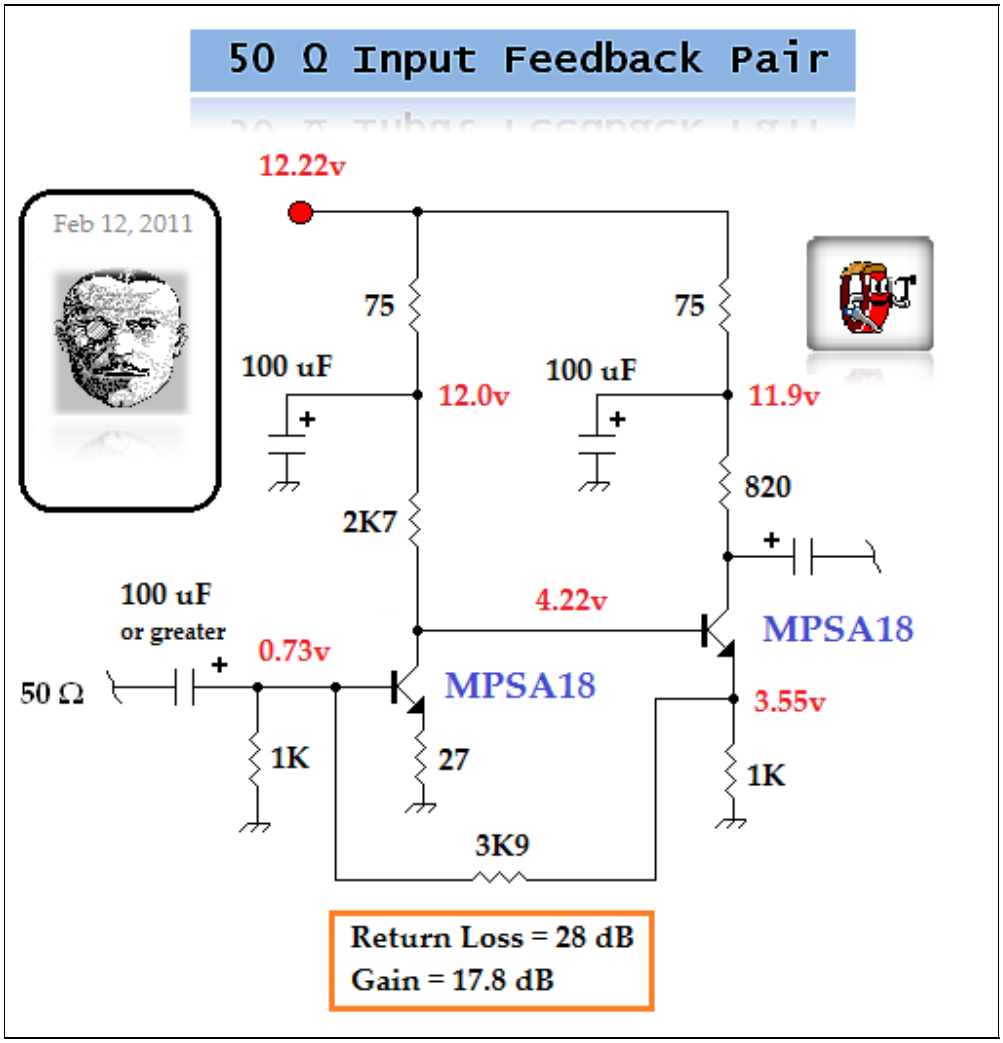
About 1/4 of my resistor collection are 1% metal film, as I don't like stocking both 5% and 1% tolerances of the same R value. I envision this amp driving a 5532 amplifier/low-pass filter built with 5% resistors with the op-amp output connected to a 1-10K volume control potentiometer.

The amp above is pure popcorn fluff, yet represents over 8 hours of reading, calculations and bench discovery. In another version, I soldered in low noise transistors (MPSA18 for example) and 1% metal film resistors to try and drop the noise figure. Yet another stage had Q1 collector bootstrapping (3 layers of feedback). I'm not sure what I'll do with these experiments, however, at least 1 of these will end up on the module page as the update to the popcorn DC receiver mainframe.

Glenn, VE7DNL built a landmark common emitter preamp with parallel 2N4401s, shunt feedback and some 1% bias resistors. I built a version of his pre-amp without the diplexer. The input return loss in my version was only 5 dB — likely the diplexer network helps establish the input match. It appears Glenn's goal was low NF and the best match and noise figure don't always correlate. I matched Q2 and Q3 as well as I could since only 6 were available. An Hfe of 220 - 240 is not uncommon for the 2N4401 — a good part.

Here's a SPICE plot of the amp performed by Wes, W7ZOI. The two plots are S21 and S11. S21 is the forward gain in dB. S11 is the input reflection coefficient, which is just the negative of the return loss. At 1 kHz, the calculated gain = just under 42 dB and the input match = ~17 dB return loss.

To improve return loss in my breadboard, I need to tweek Q2 and Q3 bias (lower the 6.04K resistor), plus add some emitter degeneration. Emitter degeneration (series feedback) would worsen the noise figure. The 100 uF emitter caps offer some series feedback and capacitor values like 1000 uF are needed if "true" AF bypass is wanted.

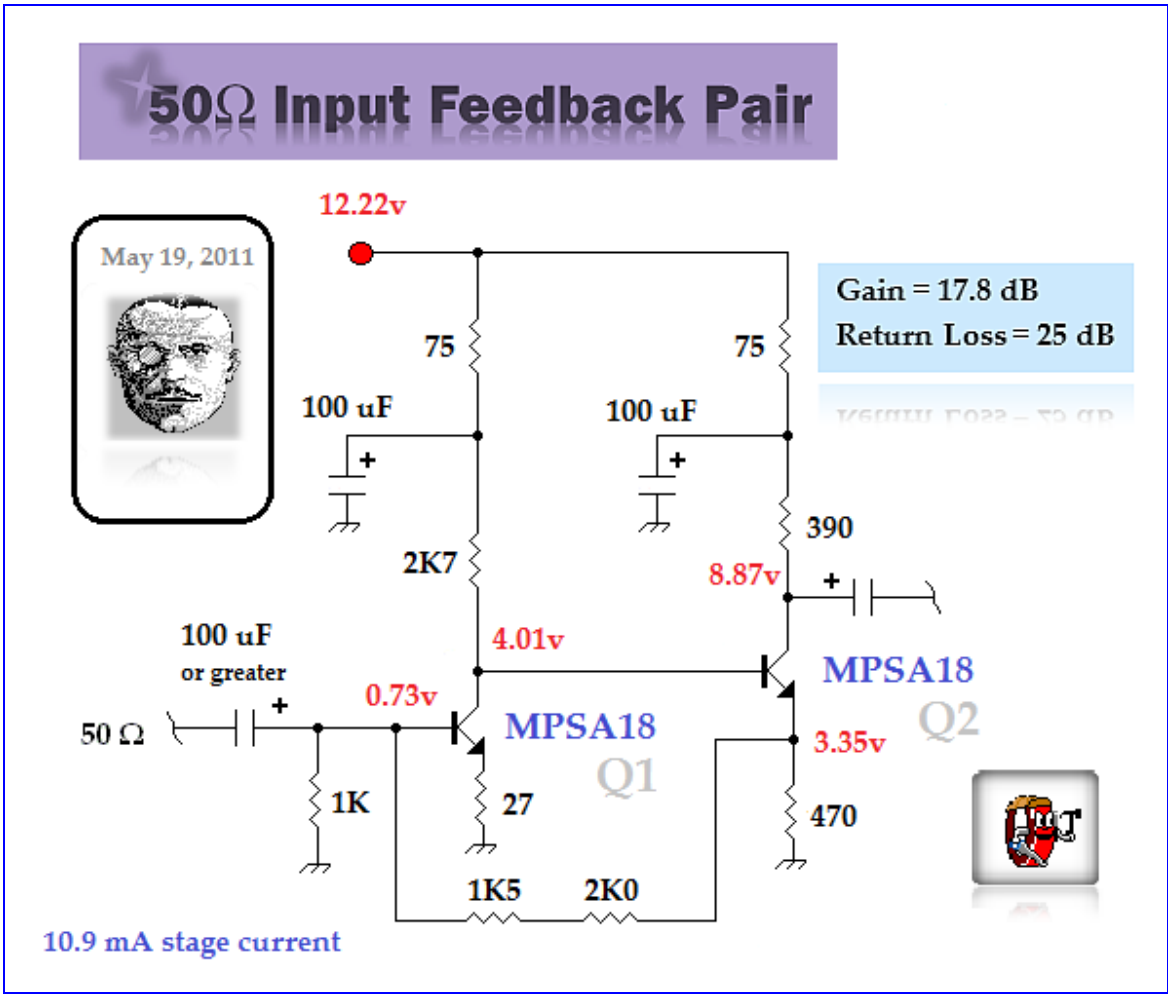


Above — Alternate experimental 50 ohm audio preamp to follow a diode ring product detector. The high impedance output enables small value output coupling capacitors for pseudo high-pass filtering. I simply added a collector resistor plus a decoupling network to Q2. The low noise MPSA18 BJT is a favorite and predictable part. The 2N3904, 2N4401 and others should work fine.

Don't even think about using an input capacitor less than 100 uF at 50 ohms — you'll trash the input Return Loss. I measured the highest Return

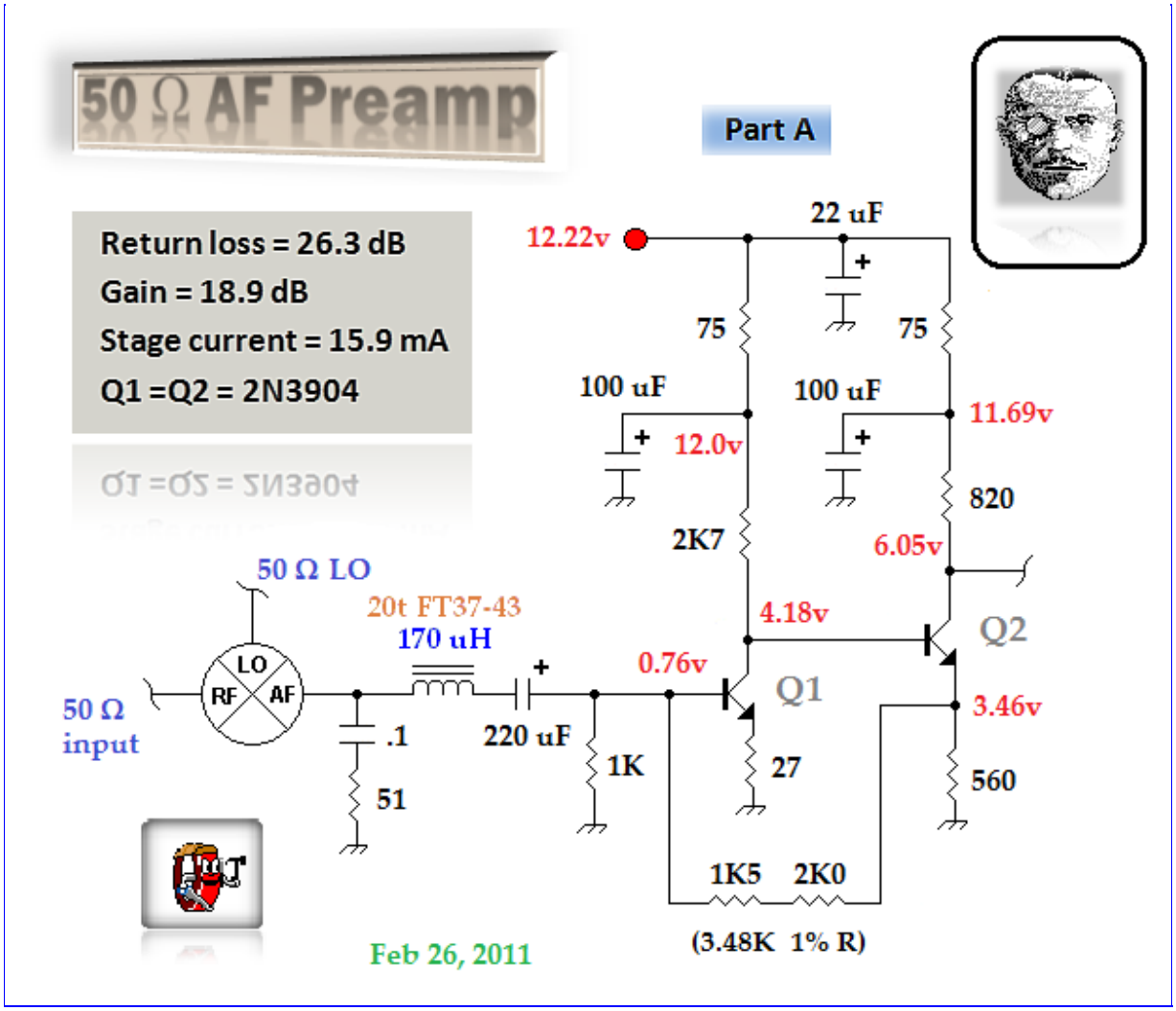
Loss with a 470 uF coupling capacitor, but the 100 uF cap shown worked almost as well.

I'm currently decreasing the (Q2) 1K emitter resistor with hopes of squeezing out a little more gain while trying to preserve strong signal handling capacity. I've found that heavy feedback is your friend.



Above — Another version; trying to increase stage gain. The Q2 collector resistor dramatically affects clean signal handling — lowest distortion occurred with a 367 ohm collector resistance. I substituted a 390 ohm collector R. This dropped the gain a little, but increasing distortion over gain is never my goal.

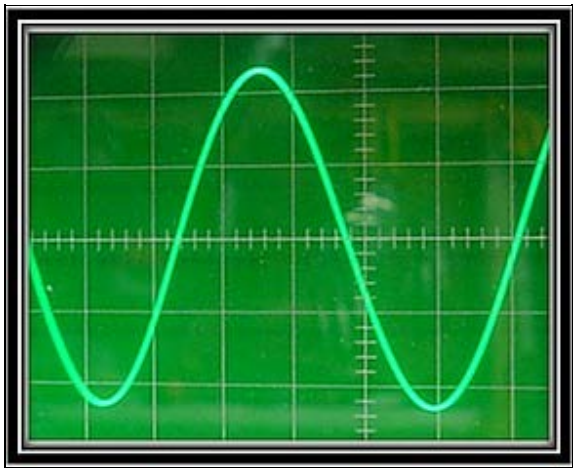
During experiments, dropping the the Q2 emitter resistor below 470 ohms made the other biasing resistors more critical, decreased return loss and invited distortion. The best return loss occurred when critically biasing with 1% resistors — I'm avoiding that!



Above — Another popcorn part experiment from February. A low-pass network terminates a diode ring mixer and drives a 50 ohm input feedback pair arranged so the Q2 collector DC voltage is approximately 1/2 VCC — this provides the DC bias for a 5532 op-amp low-pass filter arranged in the Sallen-Key topology.

The Q2 collector voltage removes the need for a separate single supply op-amp bias network and an AC coupling cap. I didn't try this circuit in a receiver; arranging the feedback network and testing the possibility of such a design drove this experiment. A perfect 1/2 VCC Q2 collector voltage might be obtained with a standard 1% value 806 ohm R. I lack this part.

Almost Popcorn Superhet



1. Receiver Concept and Macro Diagram

Not done.

Maybe a flight-of-fancy that never crosses the finish line? This page seems to repeatedly find the back burner; a good intention that never sees completion. Sadly this is normal for some of us builders — how many incomplete projects adorn your home and workshop? Still, this web page might fuel your *own* experiments.

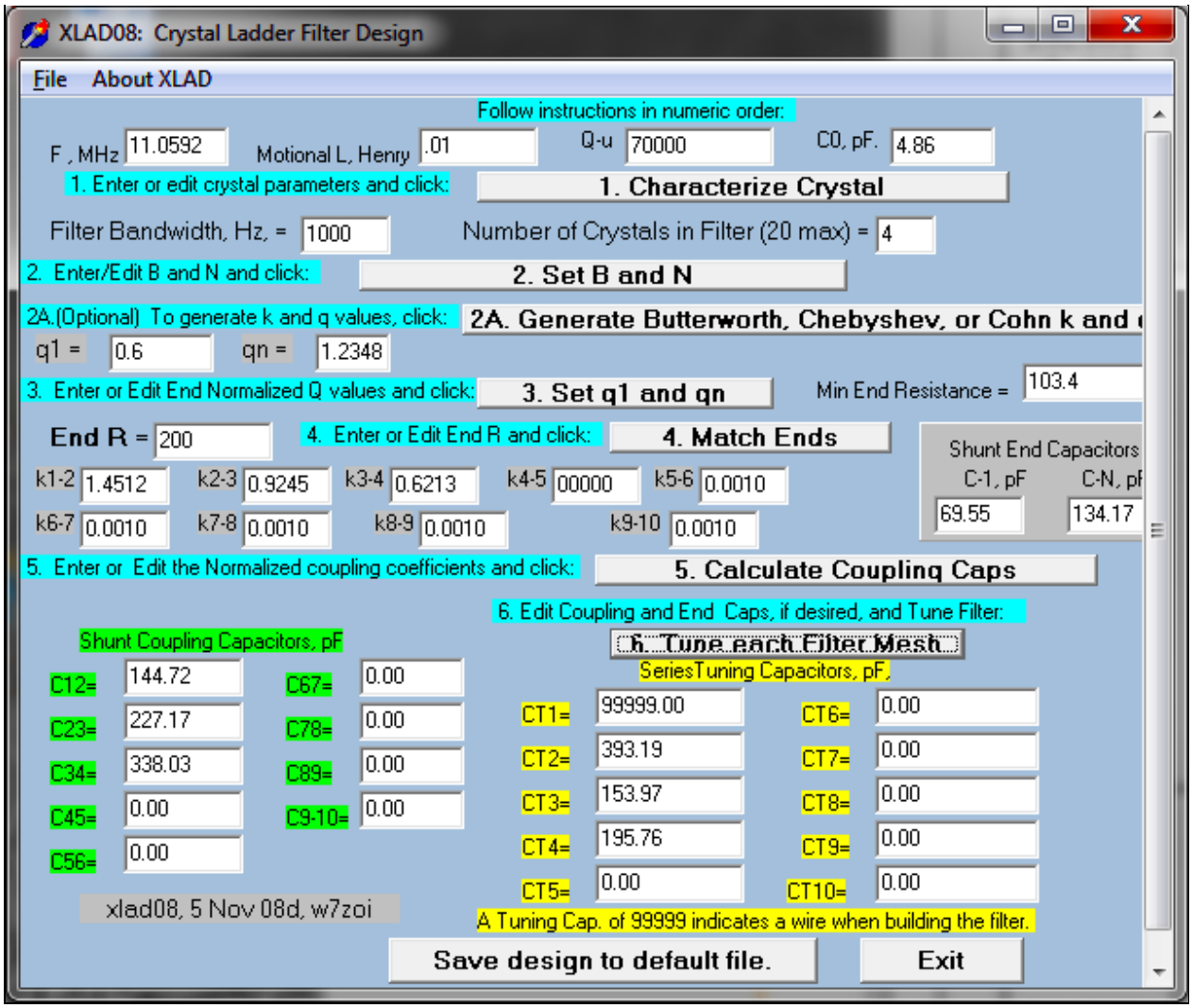
For years, I've attempted to update my Popcorn Superhet from 1998 but always seem to lose interest from numerous distractions such as VHF and UHF experiments and other non-radio stuff. So here it is so far — raw and unrefined: more experiments from my analog-centric workbench. To clarify — I love the new digital radio stuff, but that's well covered on the net by numerous code writers and those bleeding edge, frontier-pioneer types. I prefer a 50 Ω lab with all the visceral enjoyment and learning it has to offer. To each his own.

This web page remains unlisted on the main top level menu — and for good reason.

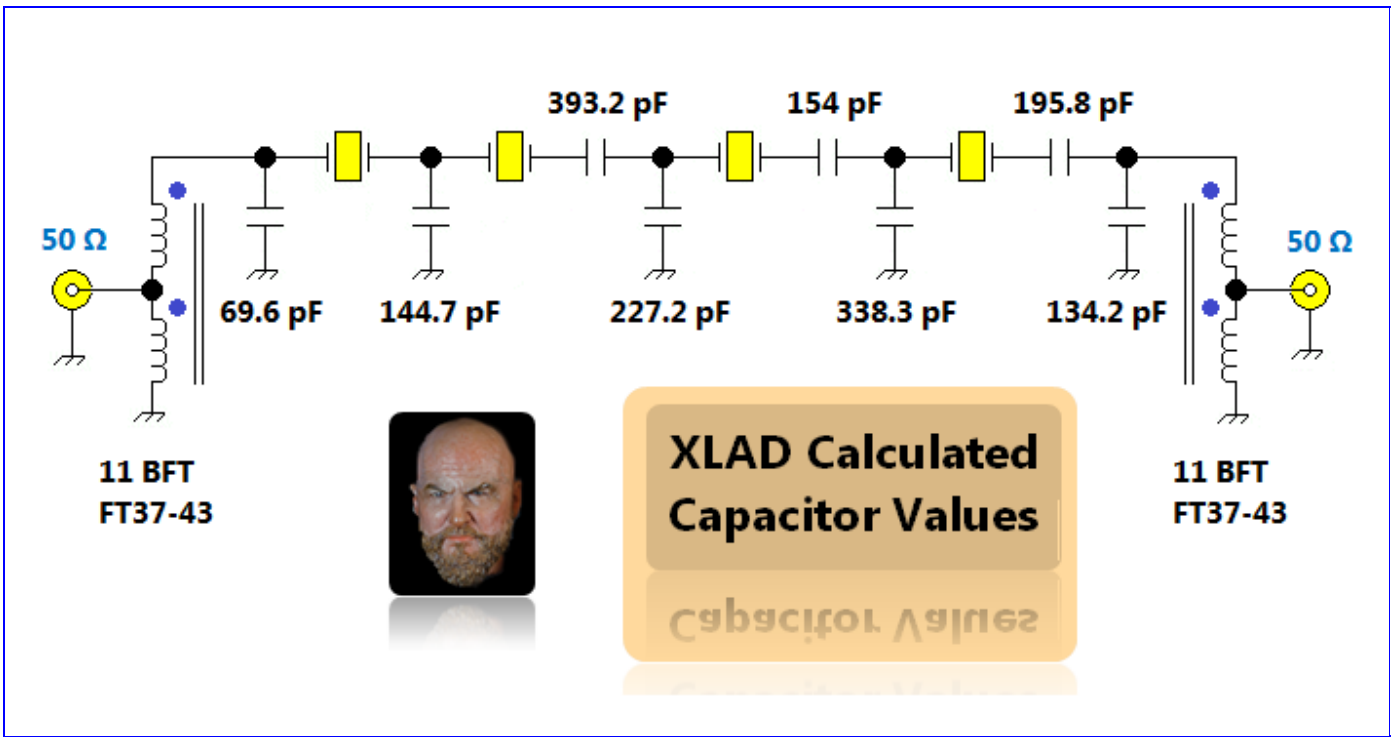
Click for some supplemental lab notes.

2. N = 4, Gaussian-to-6 dB Crystal Ladder Filter

I frequency matched four 11.0592 crystals, characterized them and took an average of those crystal parameters plus some data from Zverev and inputted this into xlad08.exe, a program that ships with EMRFD.



Above — Xlad08.exe screen shot.

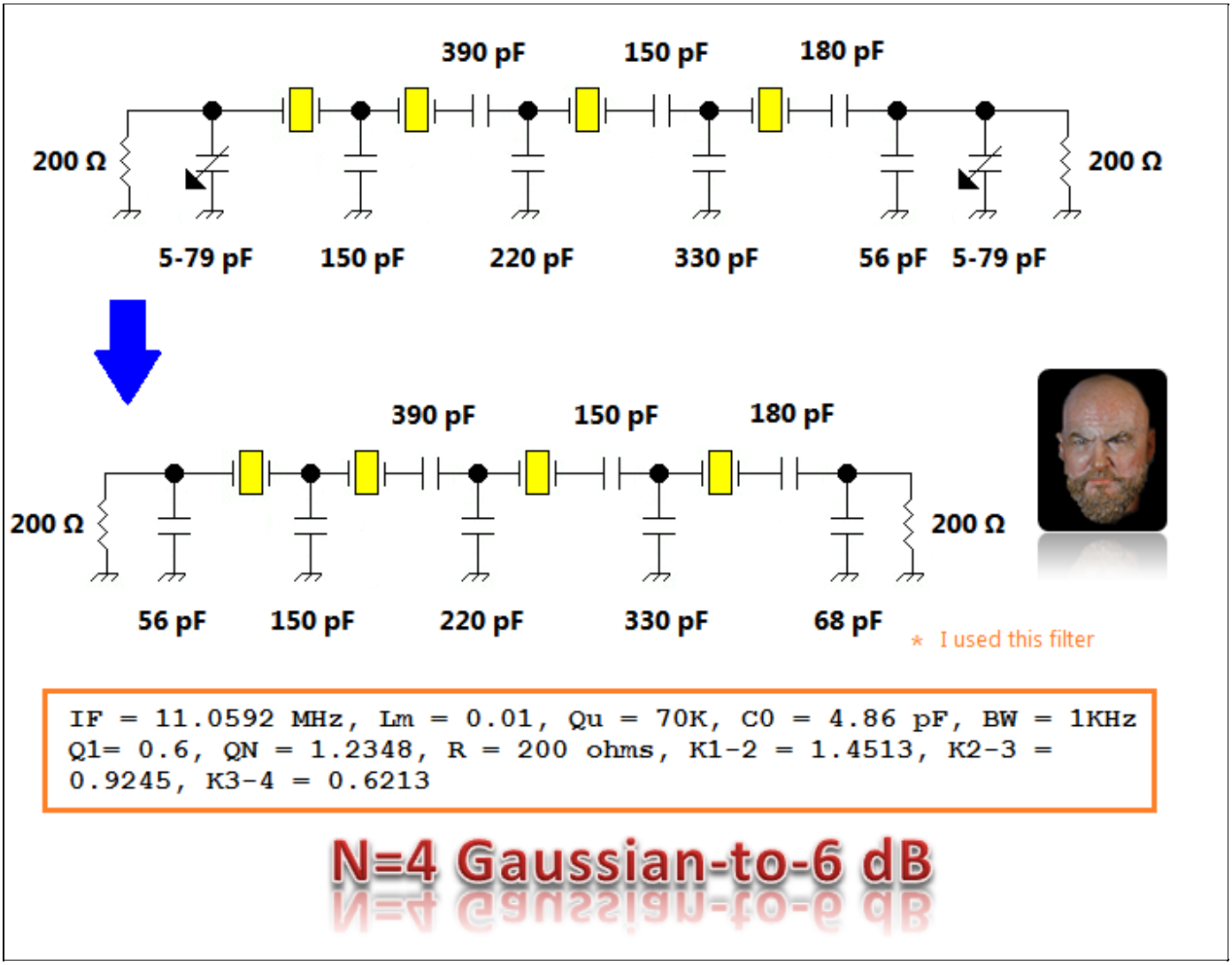


Above — The xlad filter expressed as a schematic. Click for a GPLA plot of the filter.

I built and hand tuned this crystal filter with my analog sweep system. Click for a photo of the optimized filter breadboard — I used 1 or 2 parallel capacitors to tune each capacitance to derive the best possible skirt shape. Click for a sweep of this optimized filter in my 'scope. 2 bifilar transmission line transformers provided the required 200 Ω port termination in my 50 Ω sweep system.

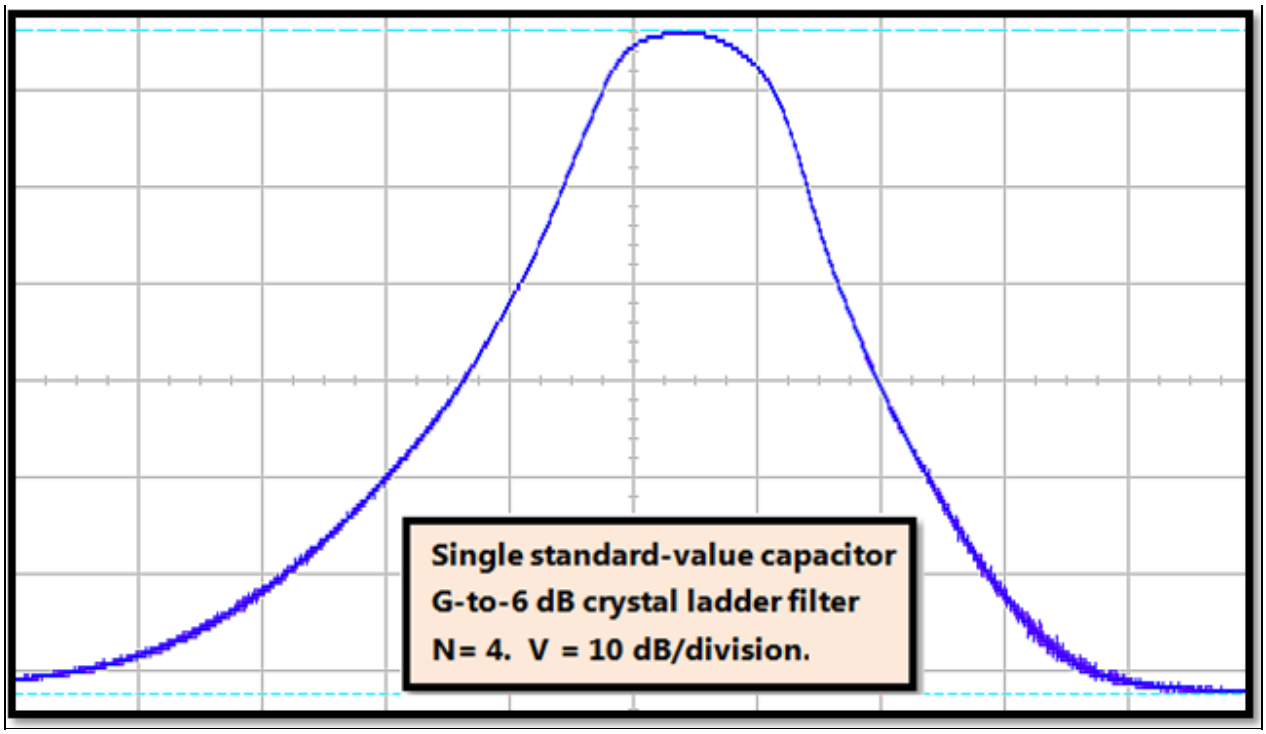
I sought a crystal filter based on the xlad08 calculated values using only 1 nearest standard value in each capacitor slot.

Transitional filters like the G-to-6 dB, are generally tolerant of part substitutions and thus I sought a filter with only 1 nearest 5% standard-value capacitor in *every* slot instead of 2 sticking 2 close in value parallel caps to make up the needed capacitance as needed.



Above —The 2-step transition to a 1 capacitor, nearest standard value in each capacitor slot. In the top diagram, I placed a trimmer cap on each end to allow filter tweaking while looking at the swept filter in my 'scope. I've learned that adjusting the parallel cap nearest to each port a simple way to tweak a G-to-6 dB filter. After setting the best looking skirt, I removed and measured the end parallel capacitance and substituted the nearest standard fixed capacitor. Manipulating the capacitors values will change filter bandwidth and return loss.

You might compare the lower filter to the xlad calculated values to see how tolerant the circuit is.



Above — A sweep of the single value capacitor version that's used in the Almost Popcorn receiver. While the 1 KHz filter skirt isn't perfect, it's reasonable. Click for the test bench breadboard. This particular filter measured 1/2 power bandwidth = 1227 KHz and the insertion loss was ~3.6 dB.

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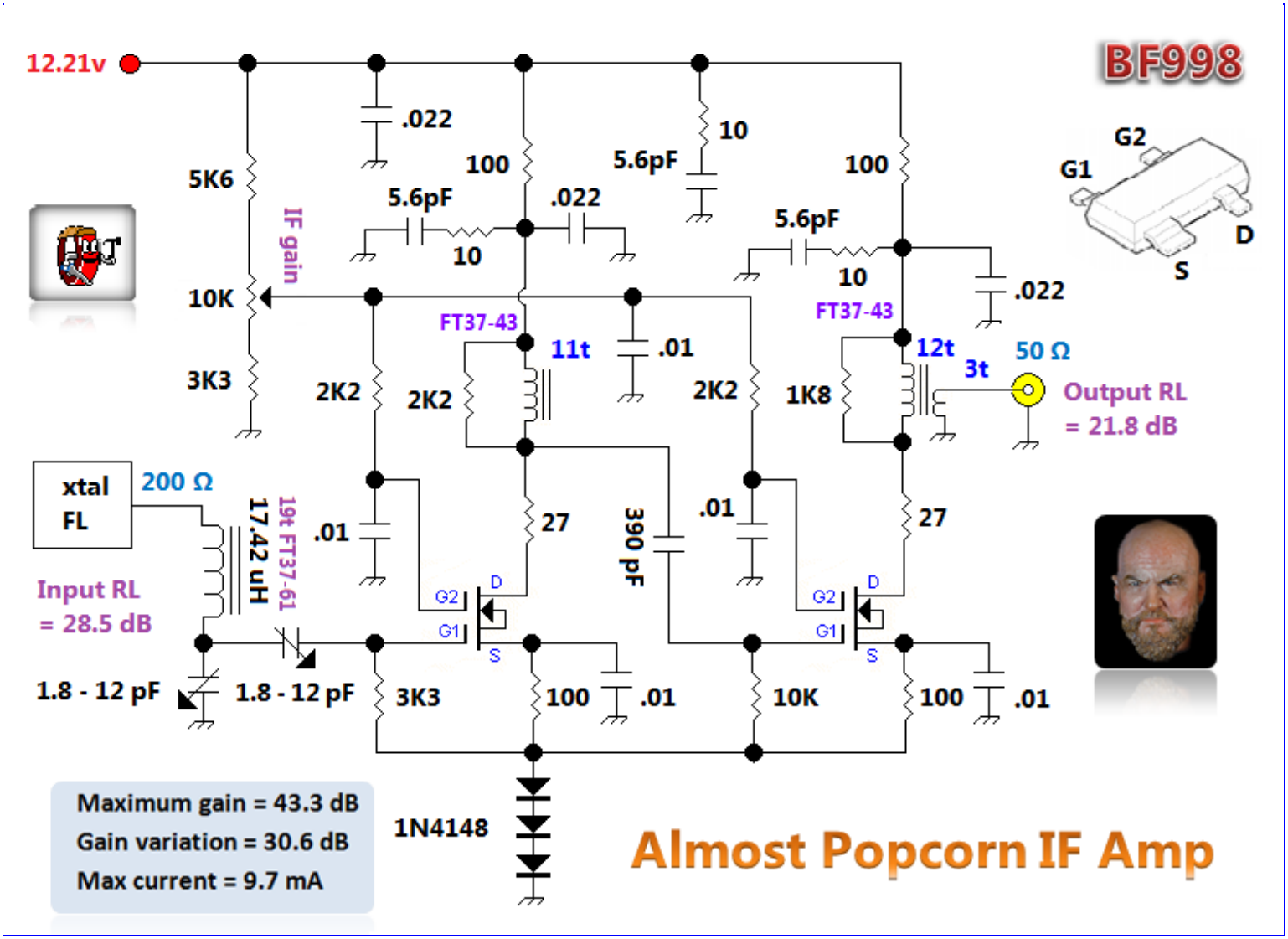
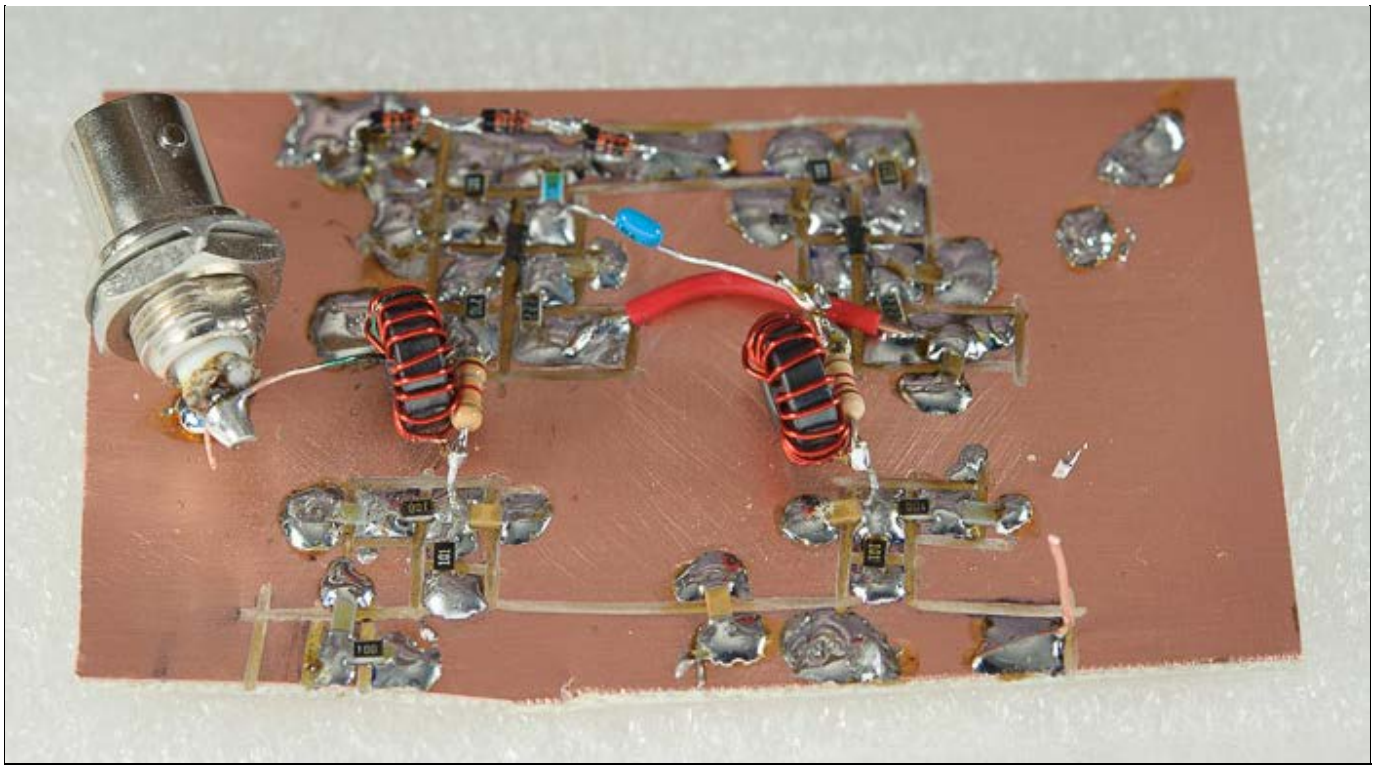
3. IF Amplifier



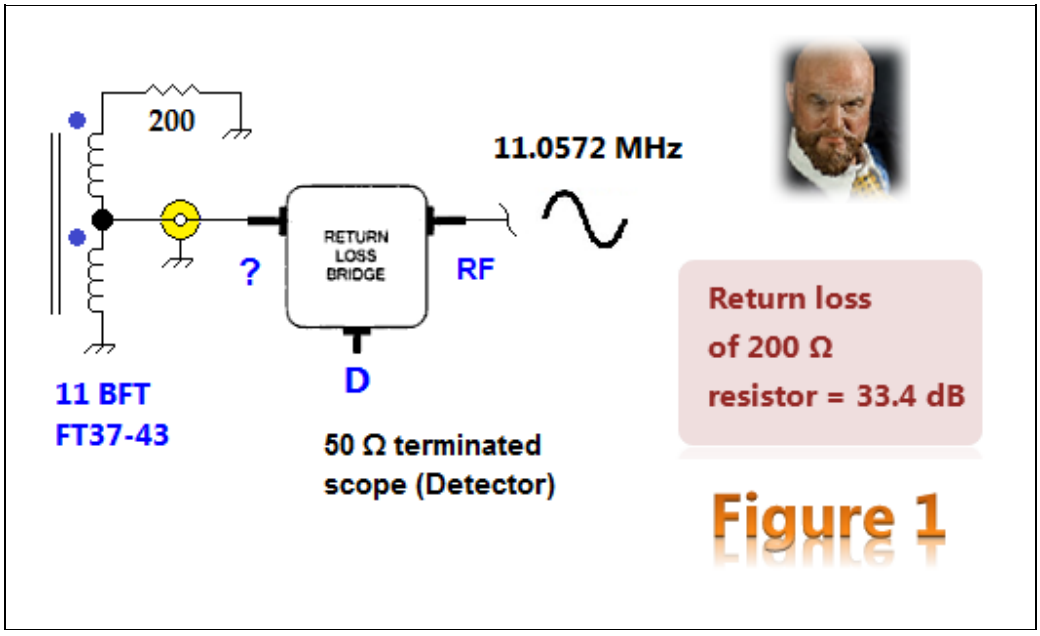
Above — Version 2 of a popcorn IF amplifer idea with BF998s. IF = 11.0572 MHz.

I have no use for AGC in my personal receivers since I listen through a speaker and *always* ride the RF gain control with my finger. Since the BF998 has gain into UHF, wideband decoupling + bypass SMT R-C networks filter each MOSFET in case of oscillations. Click and click for photos. This circuit will operate fine with just 0.022 µF bypass caps.

A network matches the 200 Ω xtal filter output impedance to the 3K3 input Z of the IF amp cascade. The L-C-C network has become a favorite matching weapon — 2 tweaks is often better than the 1 offered by a garden variety L-network. The product detector that follows this stage is a Level 7 diode ring mixer soldered onto the audio preamp board. This receiver will feature some modish circuits aimed at low noise + high fidelity but still stay "almost popcorn". Anyhow, that's my hope.



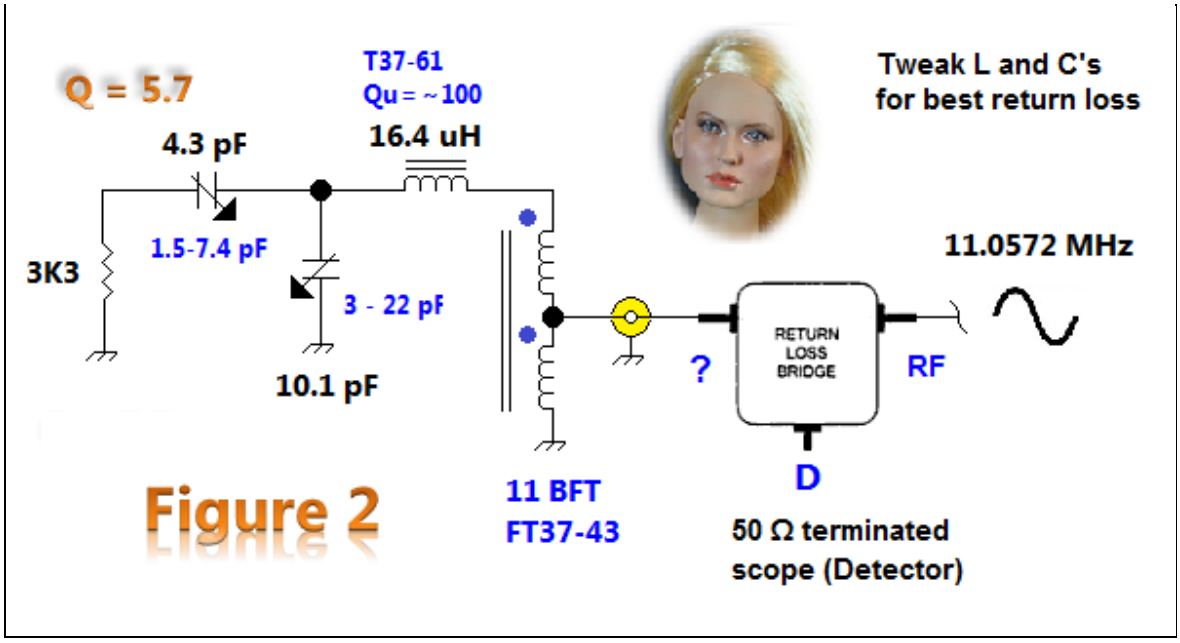
Above — A breadboard of the Popcorn IF amp sans the input network. I tested this stage using an oscillator with a 3K3 output impedance and liked what I saw. Click for a reverse angle view. After this build, I ordered some SMT versions of the 1N4148 diode. The BNC connector was only soldered on the board for testing purposes.



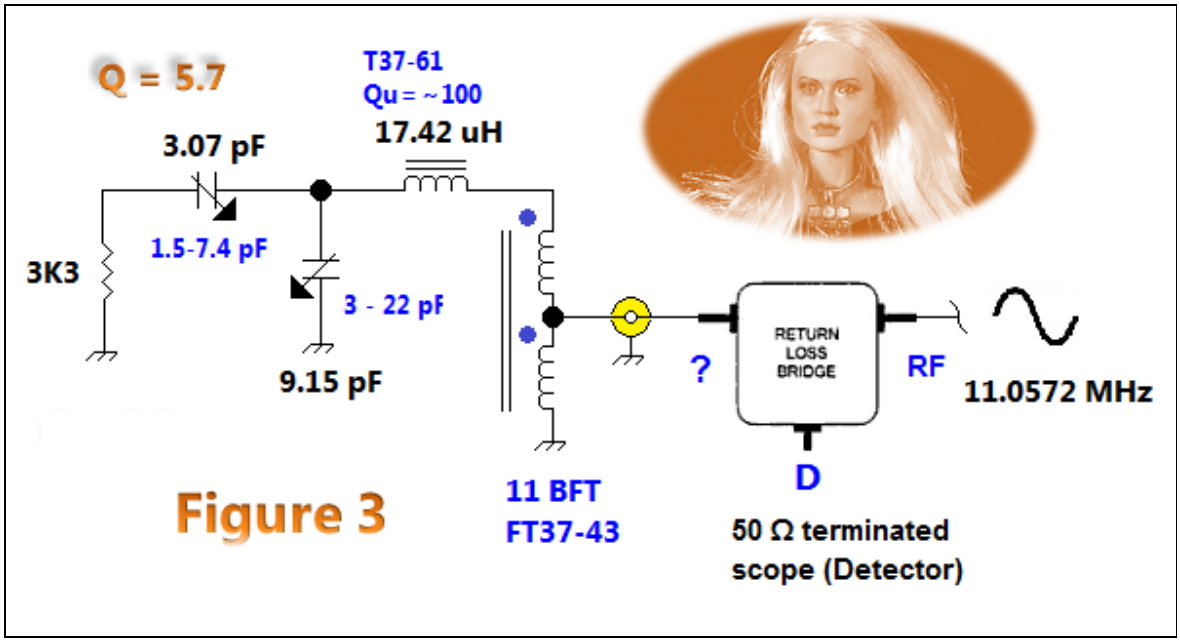
Above — Setting up to measure the return loss of the 200 Ω input Z of my popcorn IF amp. 1 difference between scratch homebrew and kit building is all the measurements you need to make to optimize your scratch receiver. A joie de vivre stems from our descriptive/exploratory analysis before and after making our breadboard. We just learn from mistakes and move forward to improve both our knowledge and gear.

I chose a 200 Ω input impedance for a reason — it's easy to apply a standard 50 Ω return loss bridge via a 4:1 transmission line transformer to establish the correct input impedance for the single frequency matching network.

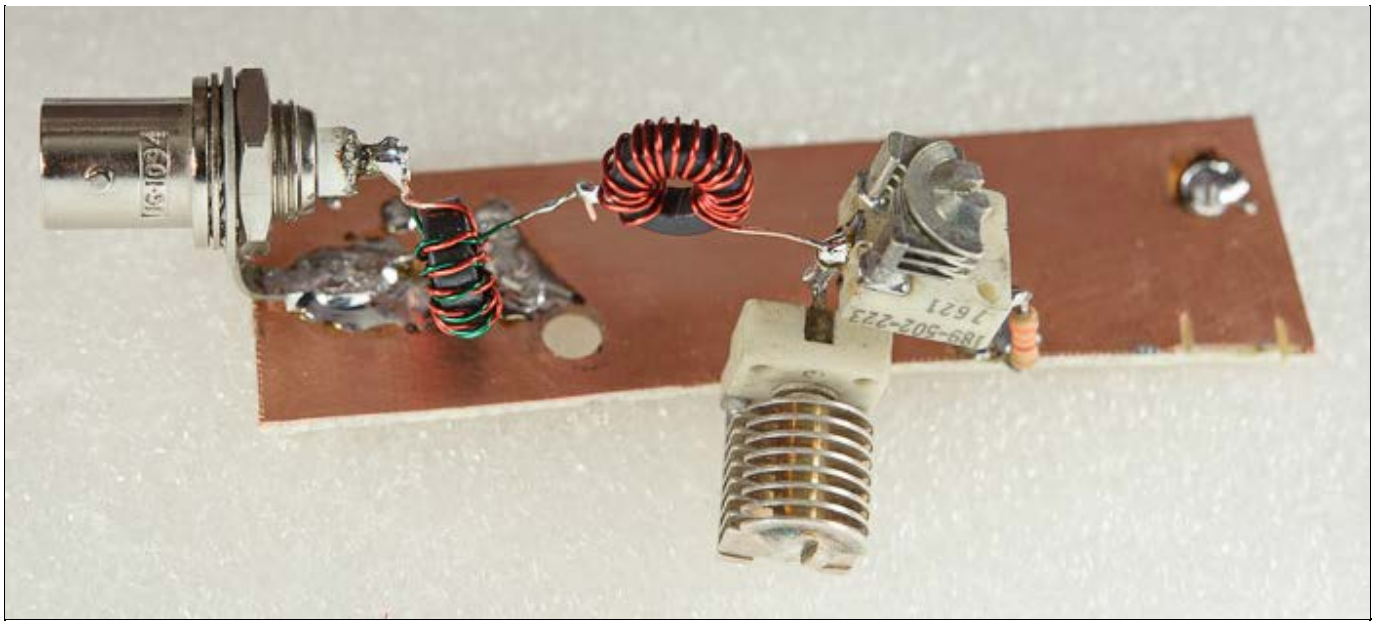
From the set-up shown above, I found that with a 200 Ω resistor, my return loss at 11.057 MHz = 33.4 dB. This signals the very best we can achieve with our network: directivity. Click for a photo of the above experiment breadboarded on a scap of copper clad board.



Above — The schematic of my 200 to 3K3 Ω matching network. I designed it with zmat08.exe, a program by Wes that I first used as a console app since the days of DOS. *Zmat08 program comes with EMRFD.* Click for a screen shot of from Zmat08.
The calculated values from Zmat08 only consider a pure resistance and not a complex impedance, so on-bench adjustments of the design L and C values are normally required and fun to do. Variable trimmer caps make tweaking the C easy.



Above — The final— *measured* schematic — While watching the peak to peak voltage on my 50 Ω terminated scope, L and both capacitors were tweaked until I found the highest return loss. L is tweaked by scrunching or expanding turns, or adding/removing turns as needed: in this case, I added a turn and expanded the windings slightly. With the L-C-C match, a return loss greater than 25 dB may arise plying good bench practices — mine measured a lovely 29.2 dB. I removed and measured the inductor plus 2 capacitors to get the values shown **black** in Figure 3.



Above — A breadboard of the Figure 3 network before I experimented to find the best L (and shortened its leads).

After I making and tweaking the Figure 3 network, I took the coil along with 2 cheaper trimmer caps and added them to my IF amplifer breadboard. While measuring the return loss of the IF strip input, I had to expand a couple of links on the inductor, but after tweaking the L and the 2 trimmer caps, I measured a 28.5 dB return loss. This ensures my crystal filter will see close to a 200 Ω termination at the IF input. I love well matched stages!

Further, I made a L network and then a L-C-C network to match the output port, however, UHF oscillations erupted. I then went back to a simple broadband transformer and measured a RL of 21.8 dB with a 12:3 turns ratio. Done.

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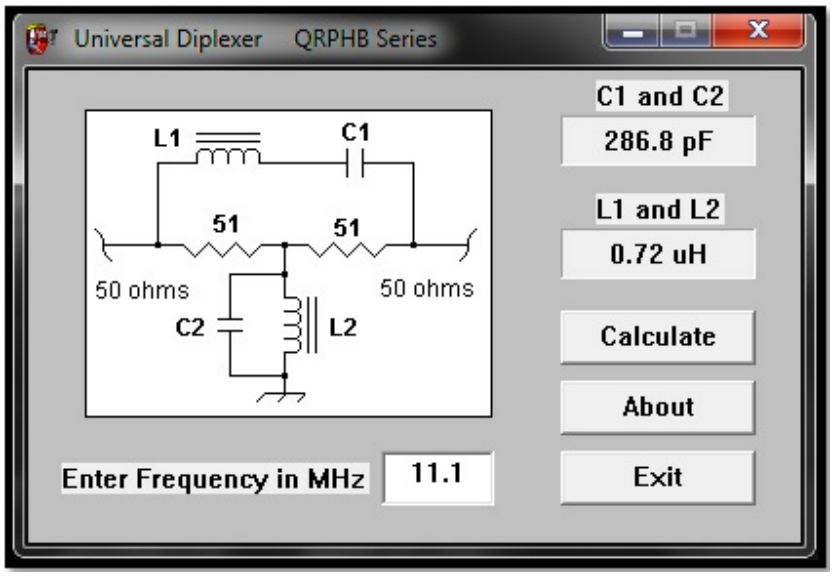
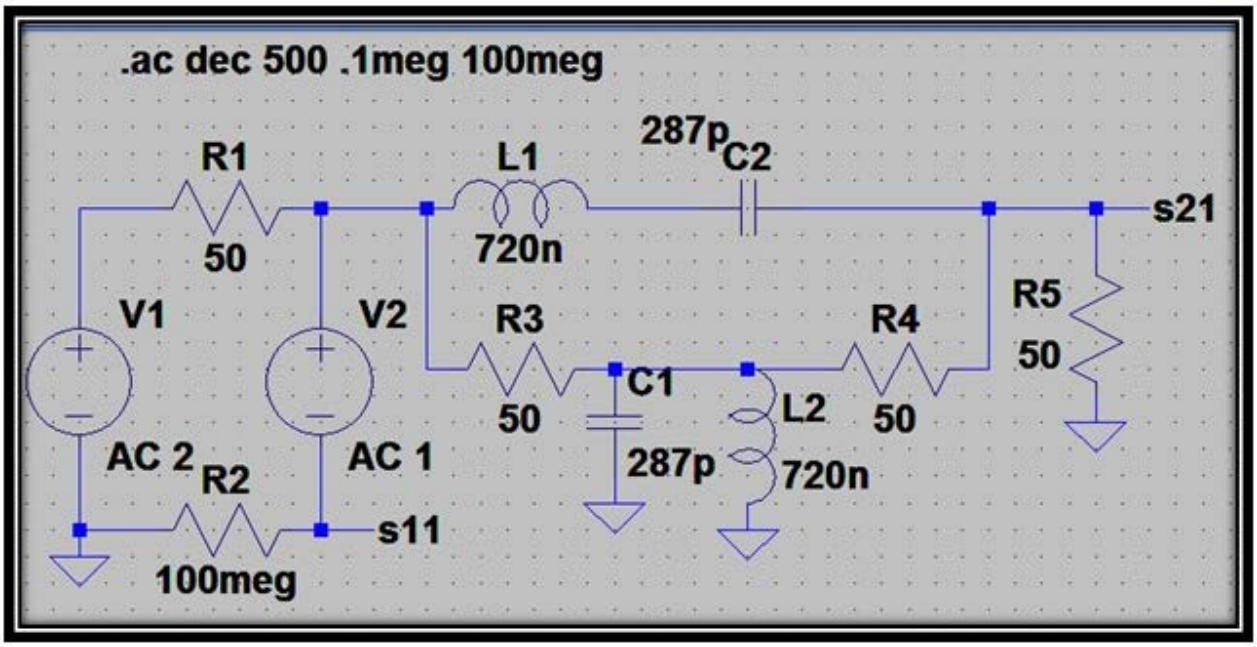
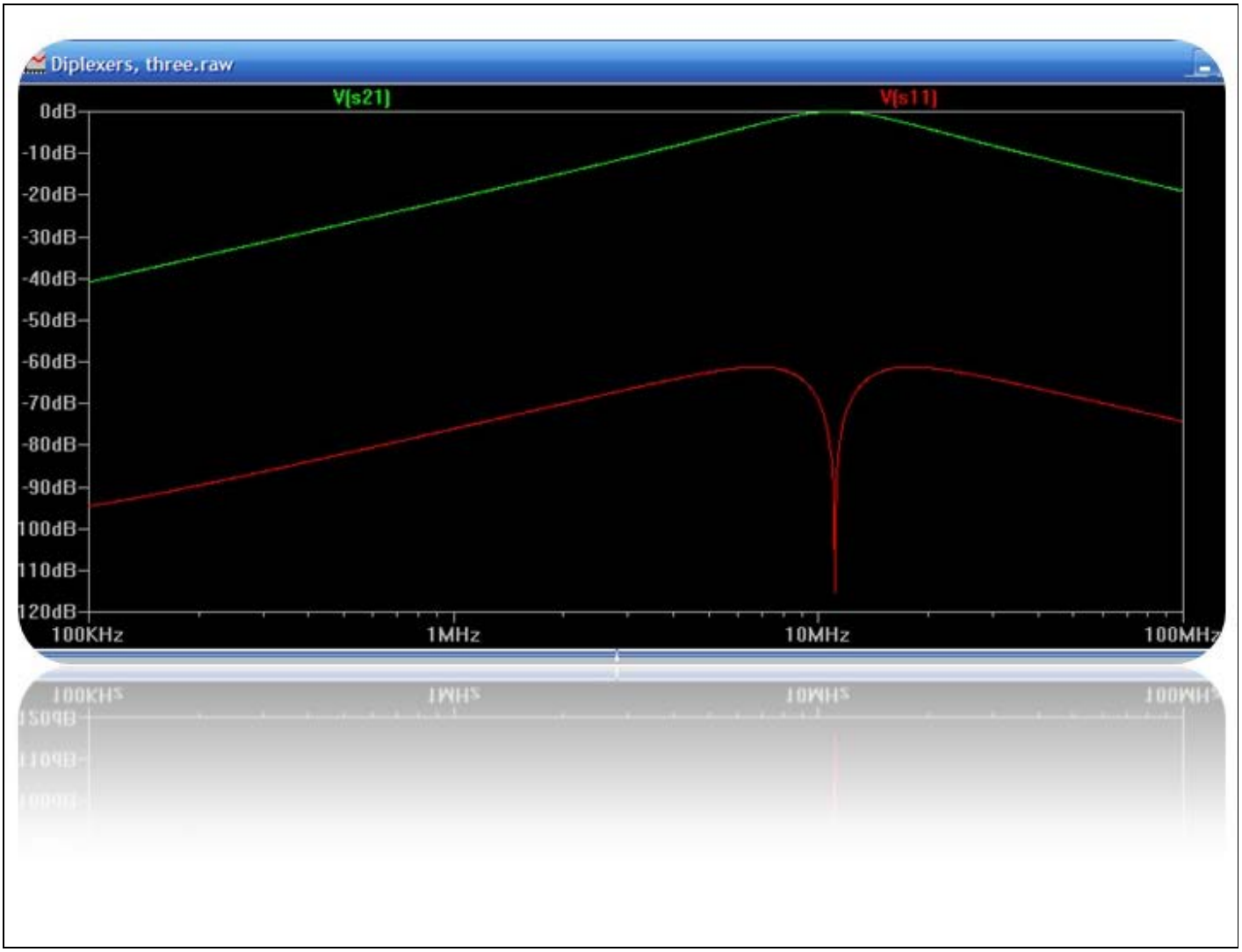
4. Mixer and Post-Mixer Amplifer



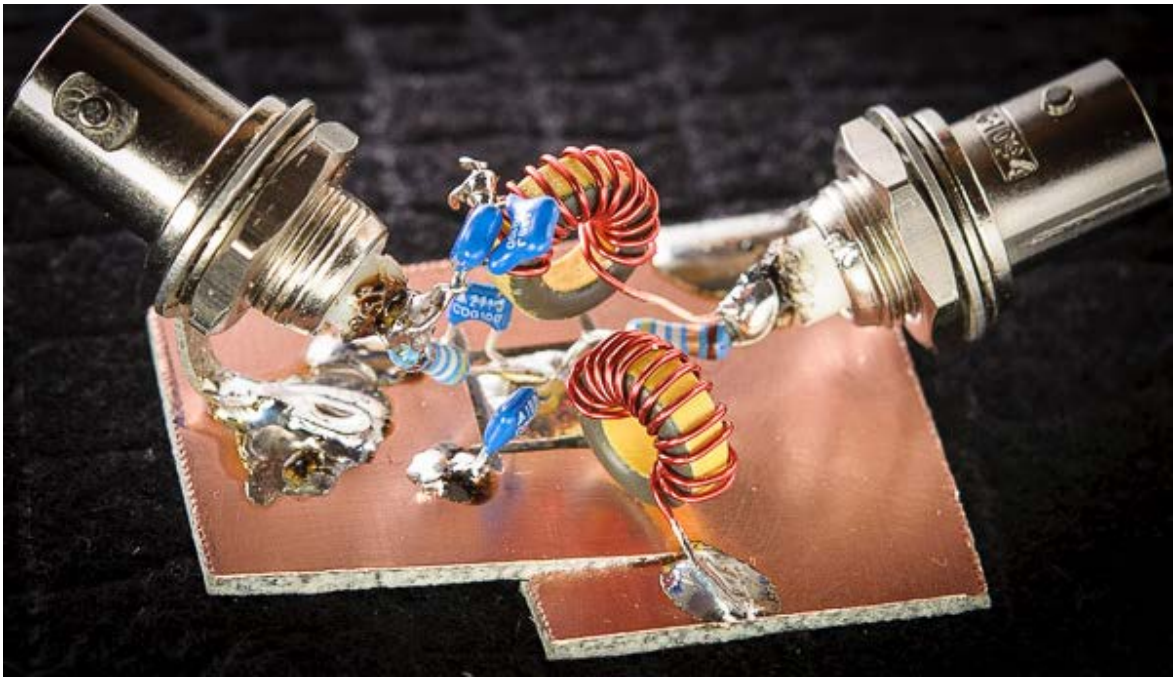
Above — I tested the W1JR (Joe Reisert) bridged-tee diplexer against another design used by Ten-Tek. I'll show the outcomes on the Almost Popcorn receiver page some day, but the W1JR better fullfills its purpose; termination from DC to daylight at the mixer IF port. I calculated the L and C values for a 11.0592 MHz IF with my Universal Diplexer application from many moons ago.



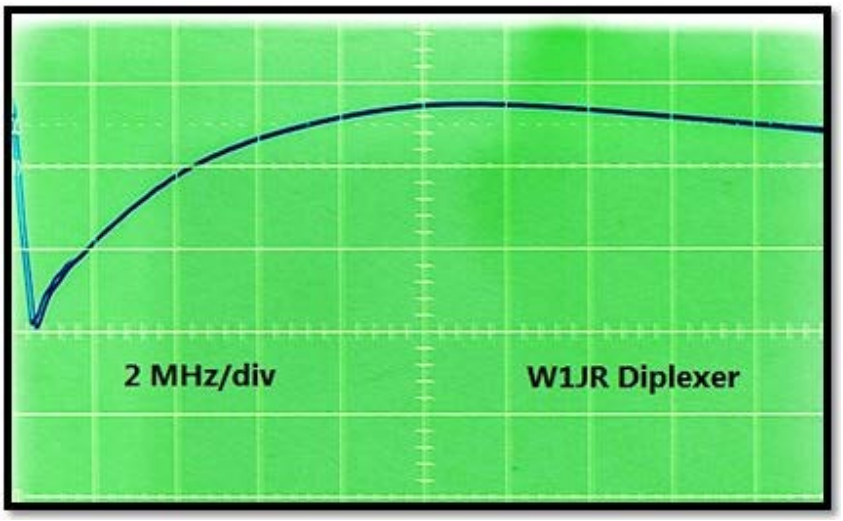
Above — SPICE provides a great way to assess a diplexer since you can plot both S21 and S11 against frequency.



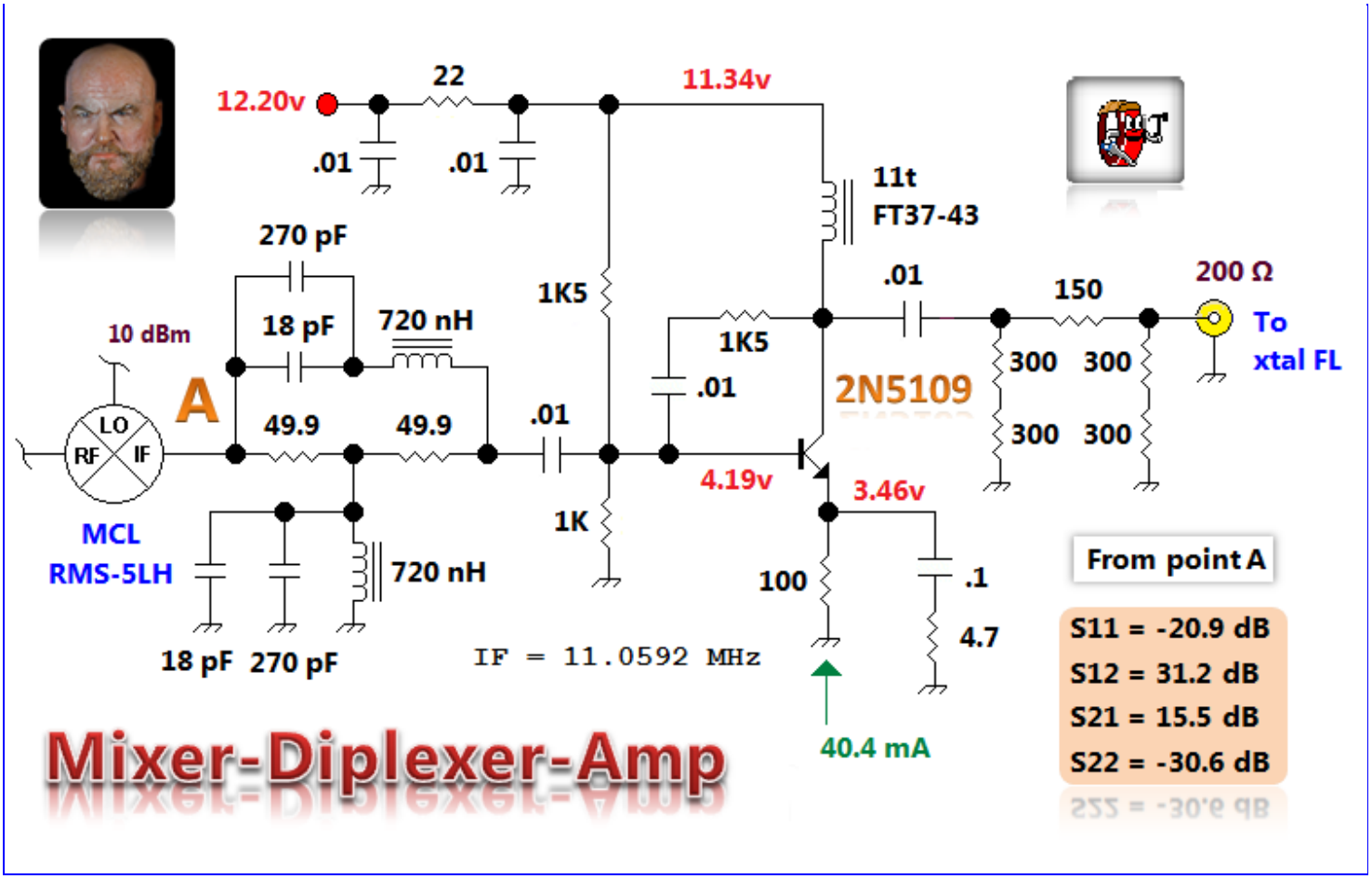
Above — A fantastic match! Thanks Joe for the design and to **Wes, W7ZOI** for his help with diplexers and the SPICE plots through many years of related experiments. Since web publishing the diplexer page, at least 500 diplexer-related emails have come in. Human translated versions of the web page exist in at least 3 languages.



Above — The evaluation breadboard of my W1JR diplexer. Those are T30-6 toroids but they look huge in this photo.



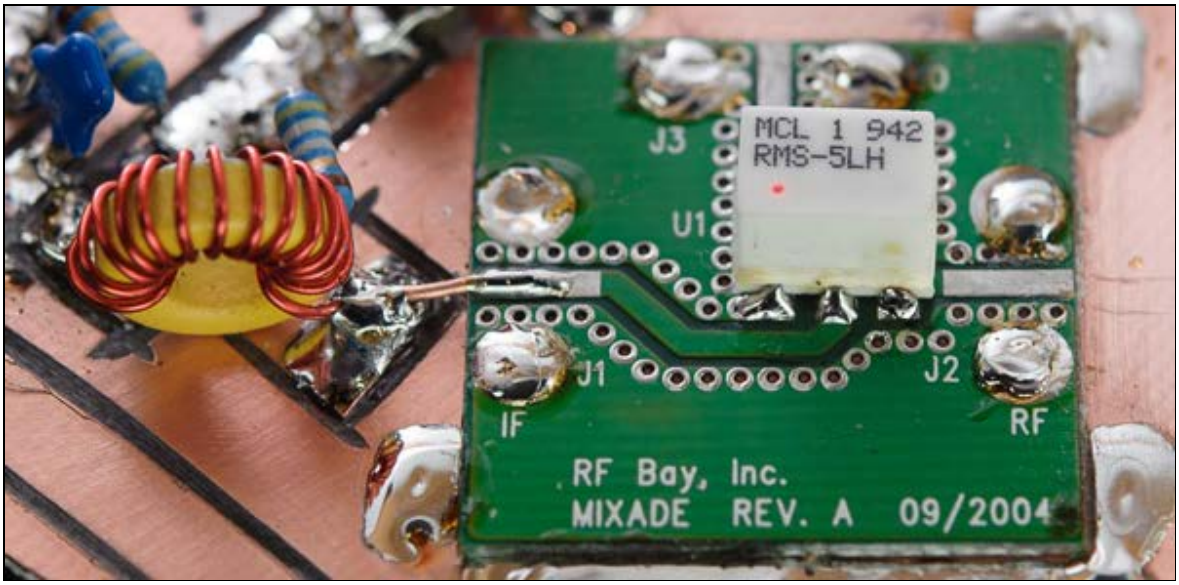
Above — Tracking generator plus spectrum analysis sweep. SPICE does a better of examining the high-pass side.



Above — A look at the diode ring mixer, diplexer and post-mixer amp from the latest version of my Almost Popcorn Receiver. I tested the diplexer and RF amp with pad as one unit (see the S-values).

It did not make sense to use a transmission line transformer to get 50 Ω on the amp collector and then use another transformer to build it back up to the needed 200 Ω impedance for the crystal filter, so I just employed a single choke driving a 200 Ω Z attenuator pad.
S parameters are 50 Ω values, thus S22 = the 200 Ω transformed into 50 Ω. After terminating the stages 200 Ω output with a 4:1 Z transmission line transformer, I tested it like any other 50 Ω circuit andmeasured S22 as -30.6 dB — my xtal filter input Z is well established.

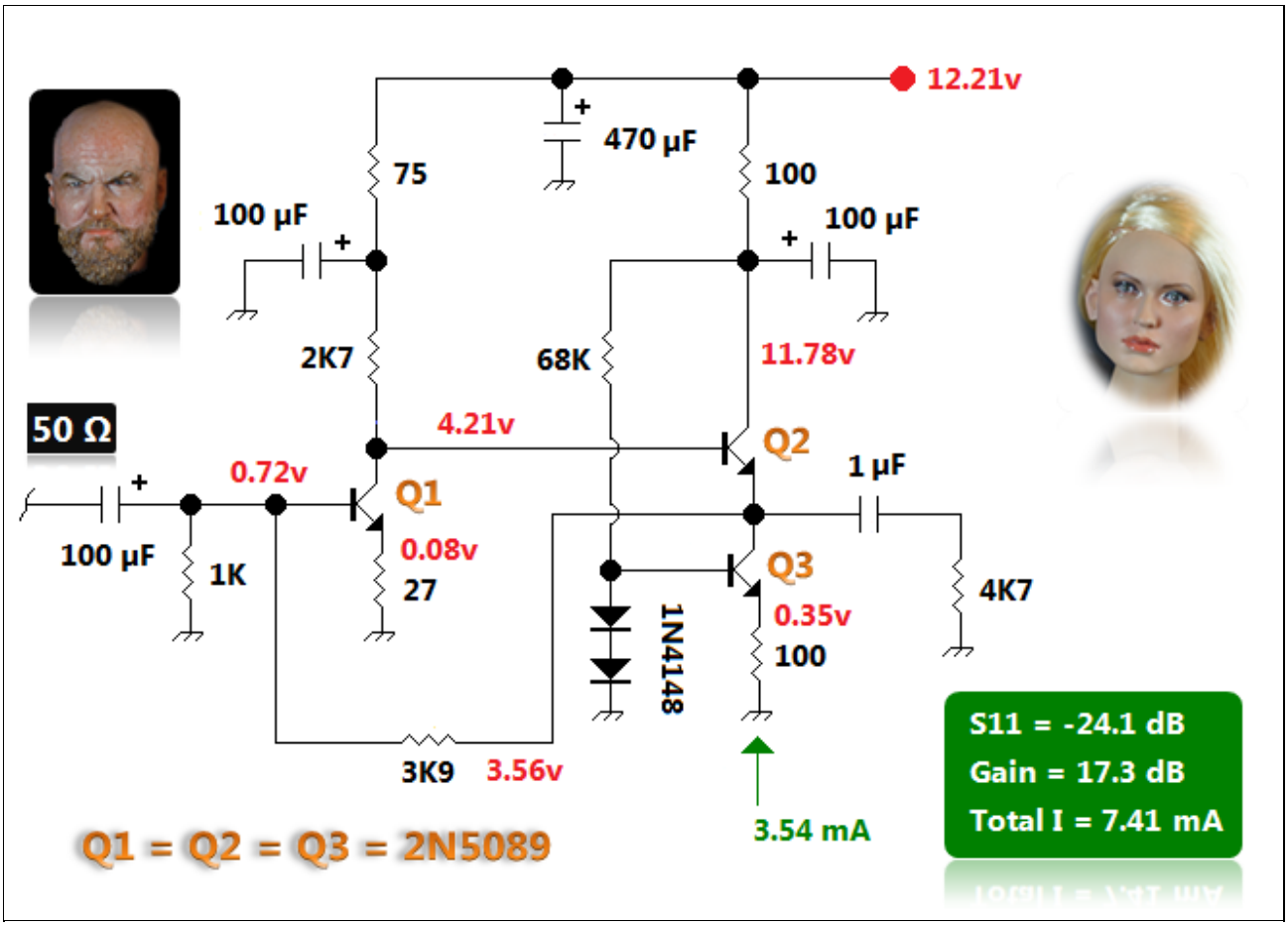
Even with the 6 dB *(200 Ω)* attenuator pad and losses from the diplexer, the stage gain from point A = 15.5 dB. Click for to view a bigger picture of the breadboard. Losses from a diode ring mixer are typically 5 - 7 dB.



Above — Mixer and diplexer assembly. The RMS-5LH takes 10 dBm LO drive.

5. Product Detector and Post Product Detector AF Amplifer

I kept the diode ring for a product detector: an SBL-1 driven with a xtal oscillator (BFO).



Above — My post-product detector amplifer: a common emitter/ emitter follower pair arranged for a 50 Ω input impedance with shunt and series feedback. Getting the biggest, clean sine wave in and out posed my only goal — the 3.54 mA current source on the follower helps that cause. The input Z tested quite insensitive to output load variations and the (S11) return loss = 24.1 dB.

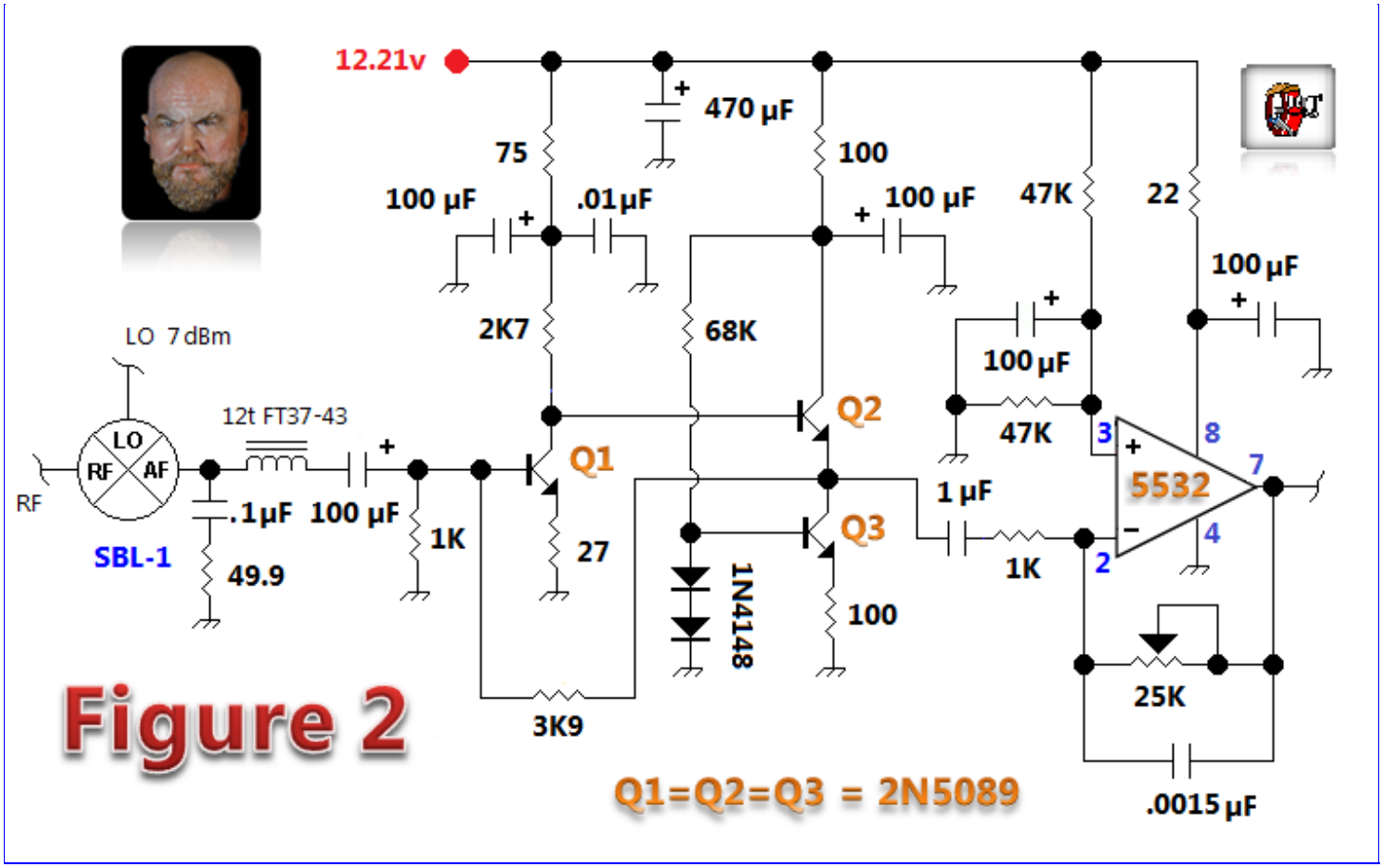
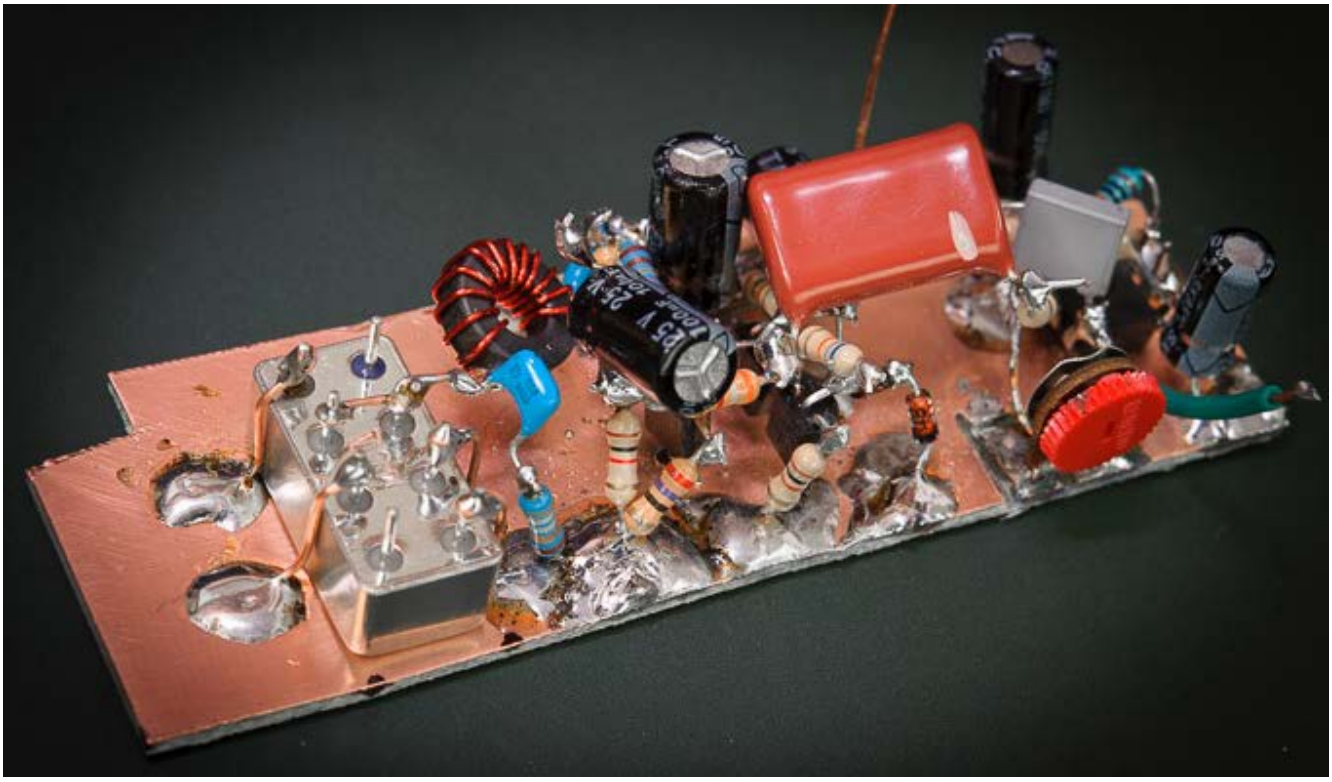


Figure 2

Above — The "Full Monty product detector to AF preamp". A MCL SBL-1 diode ring mixer drives the 50 Ω input Z feedback pair and finally a 5532 op-amp with adjustable gain. The 25K gain trimmer pot allows the builder to accommodate factors such as whether the stage is a DC receiver, or an RF-AF block for a superheterodyne receiver like in this radio. I'll set the gain as low as possible and hope to build up my AC voltage with active, gain producing low-pass filter op-amp stages.

The output t pin 7 is directly coupled to an op-amp AF filter. With software we can easily design a myriad of audio filters: band-pass, low-pass, high-pass, all-pass etc. I only run low-pass filters in this slot and even here we must decide whether to make our reponse Gaussian to 6 dB (or 12 dB), Bessel, Butterworth, or a Chebyshev with some degree of ripple. I'm testing out a few designs with an eye towards simplicity and 10% parts.

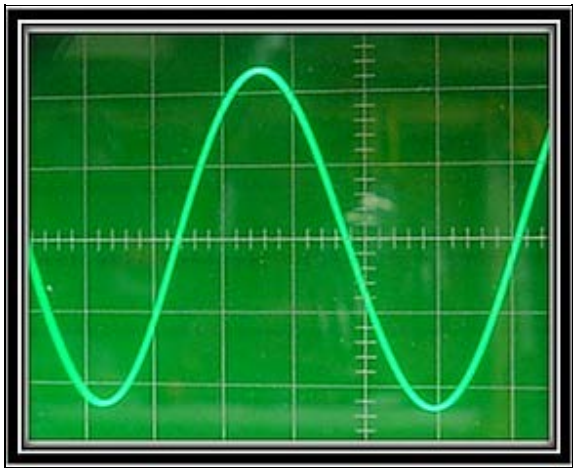
Spurred on by curiosity, with my ears as the main evaluation tool; I currently have a 5532 paper design to deliver 20 dB gain with a 500 Hz pass-band, Gaussian to 12 dB low-pass response. Click for schematic, I chose a 500 Hz cut off frequency since the skirt shape by definition is subtle and the Q low. This trick should better roll off the AF above 1 KHz a little faster without the ugly group delay associated with steeper filters employing more poles. We shall see!



Above — The compact breadboard of Figure 2.

Audio filters and AF power amp are incomplete.....

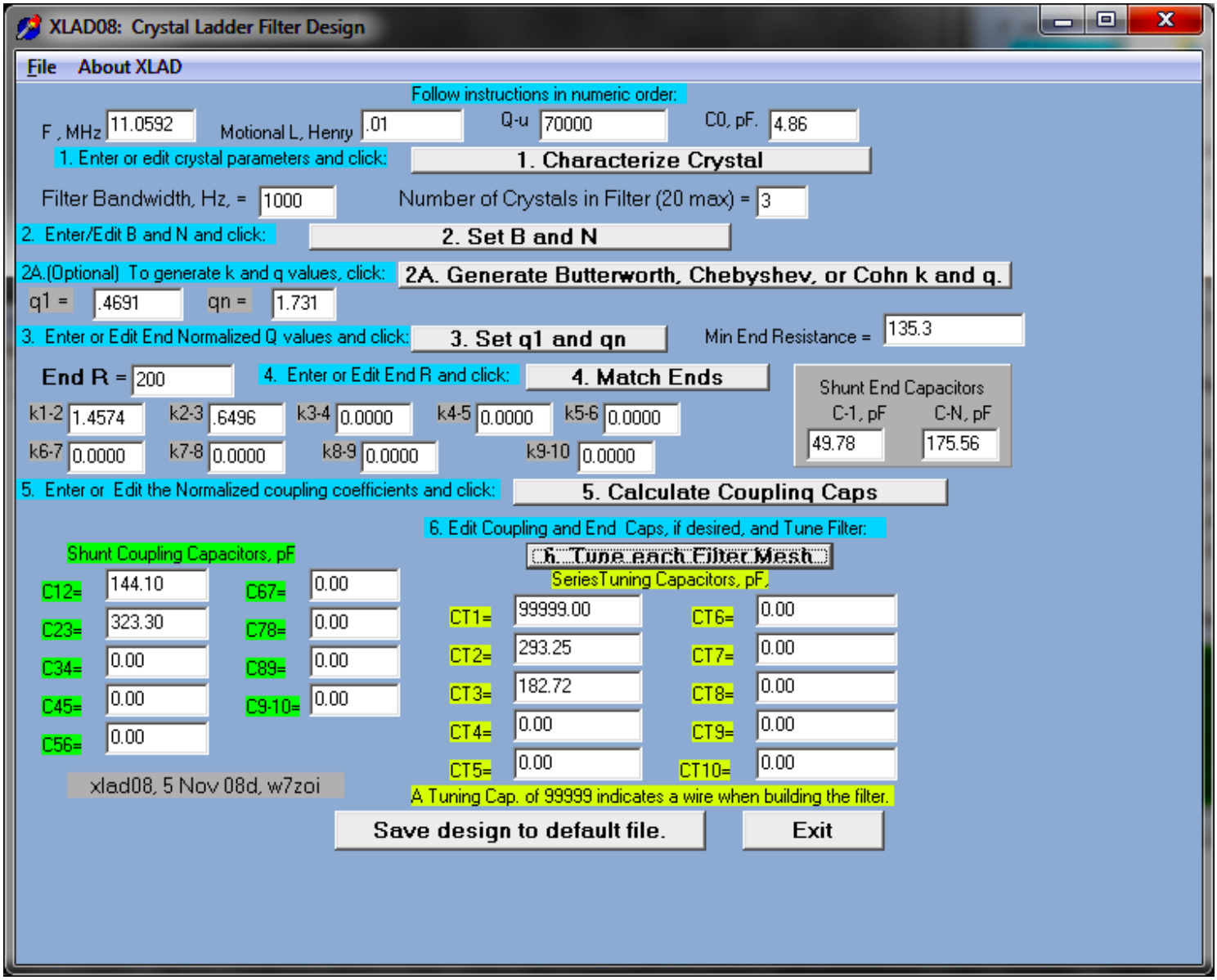
Almost Popcorn Superhet — Supplemental Lab Notes



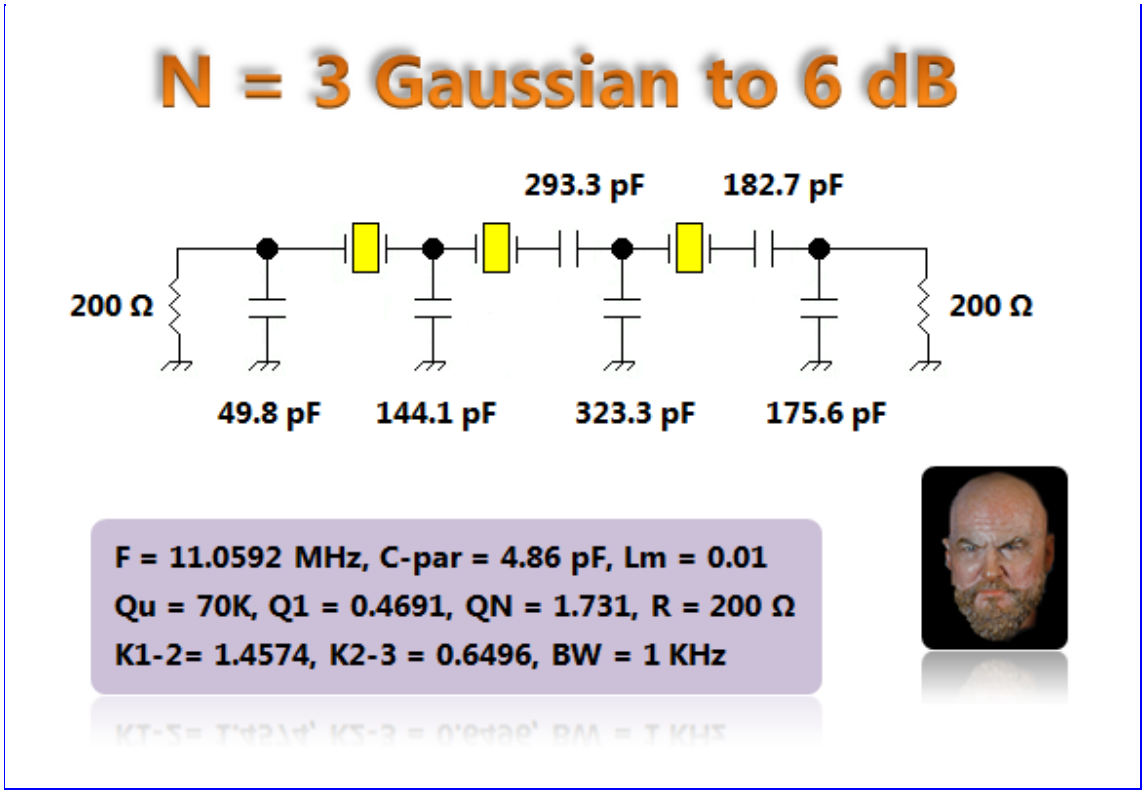
These lab notes supplement the Almost Popcorn Superhet Receiver Web page which has remained incomplete for greater than 10 years. I'm sorry for the rough writing and procrastination.

N = 3 Gaussian-to-6 dB Crystal Ladder Filter

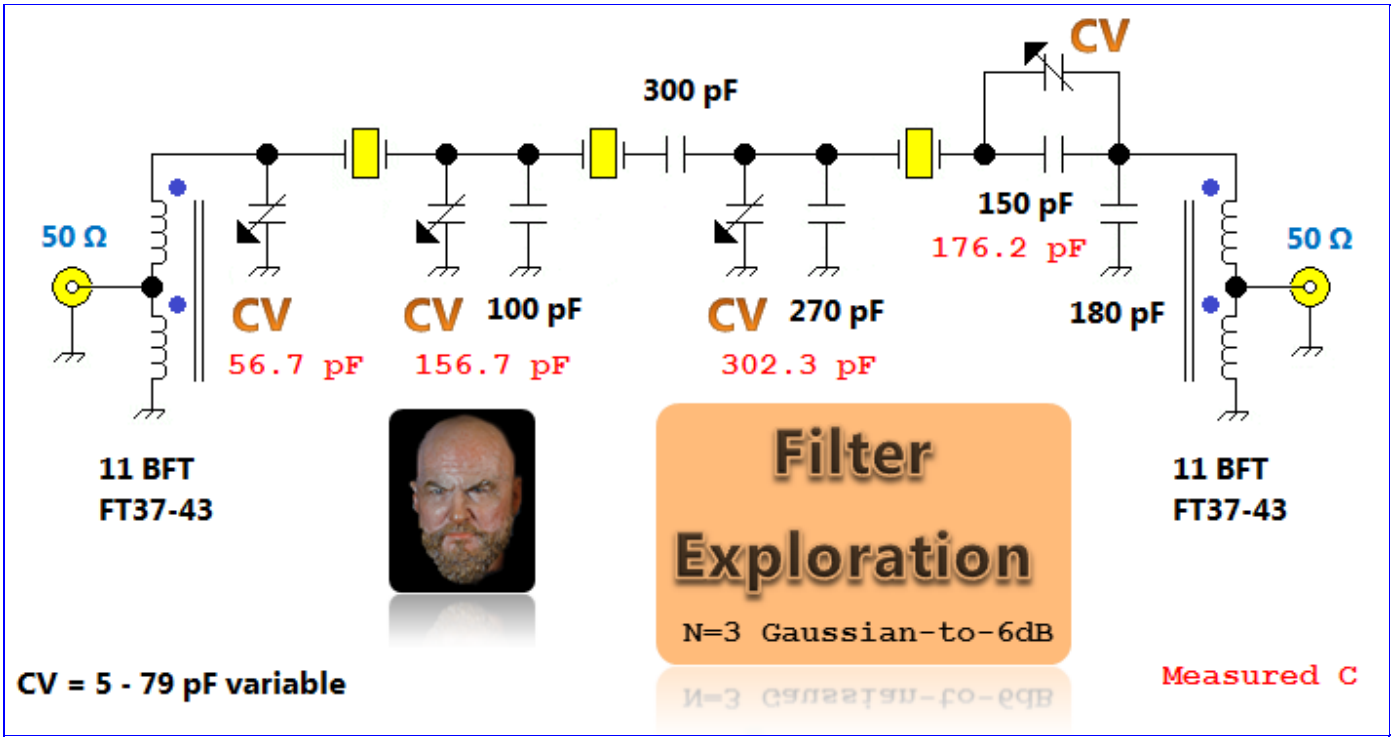
I spent a pleasant Saturday afternoon designing, building and testing Gaussian-to-6 dB crystal filters. I inputted some xtal parameters + constants from Zverev into Xlad08.exe from EMRFD. Here's the N = 3 version



Above — A screen capture from Xlad. This app allows skilled builders to depart from the typical Cohn (Min-Loss) filter and tune filter poles to derive beautiful skirt shapes and low insertion loss if wanted. I'm not a great crystal filter designer, but I've seen work from friends who have a knack for it: imagine making a N= 8 to 10 crystal CW filter with just ~1 dB insertion loss like these builders! Click for the GPLA simulation.



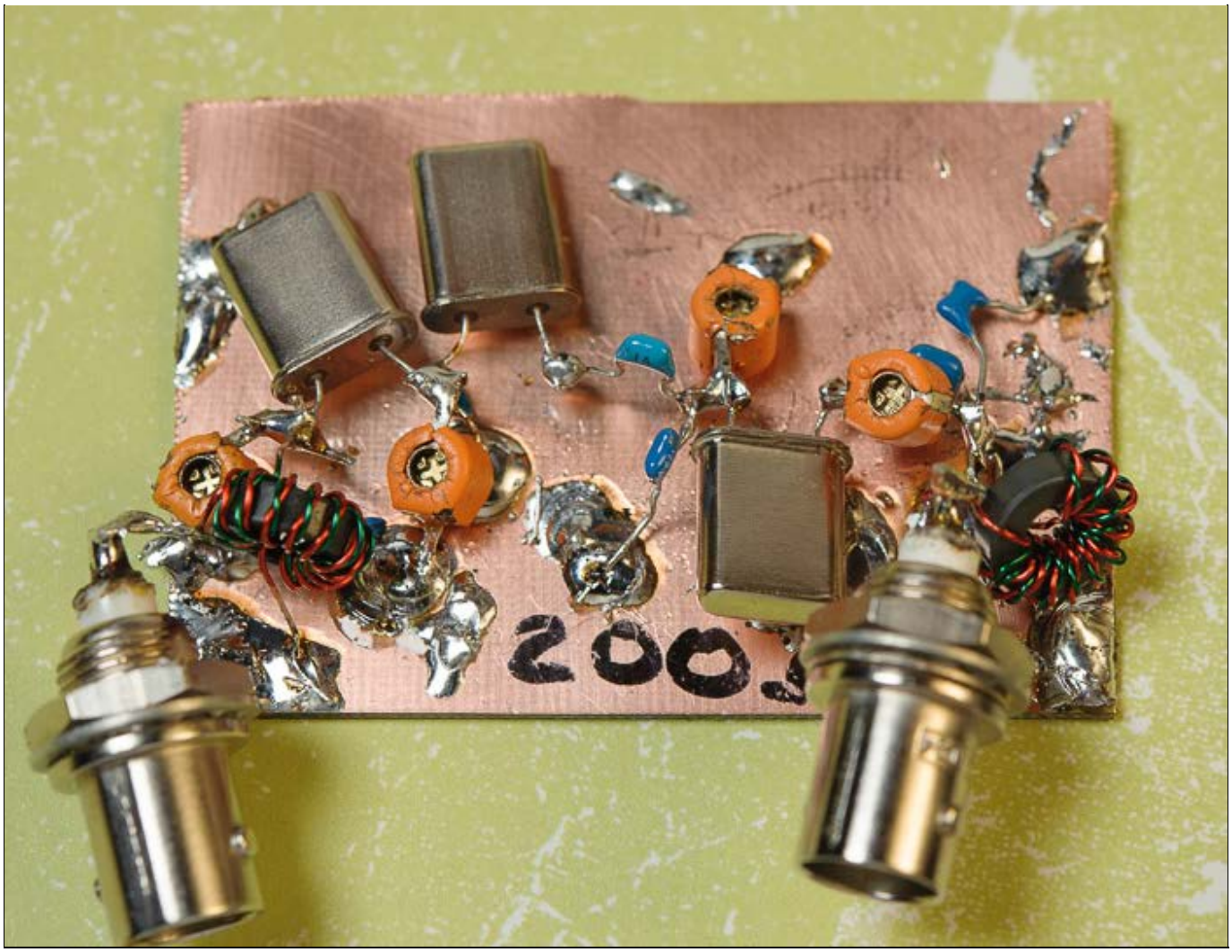
Above — Gaussian-to-6 dB xtal filer schematic where N = 3.



Above — Circuit to allow evaluation of the filter in my 50 Ω sweep system. Transmission transformers provide the needed 200 Ω filter termination. I inserted 4 trimmer caps to permit tweaking of the filter under test to establish the very best filter shape. With trimmer capacitor adjustment, I was able to trash the filter or peak it well. Click for 3 different swept bandwidths: Sweep-1 Sweep-2 Sweep-3.

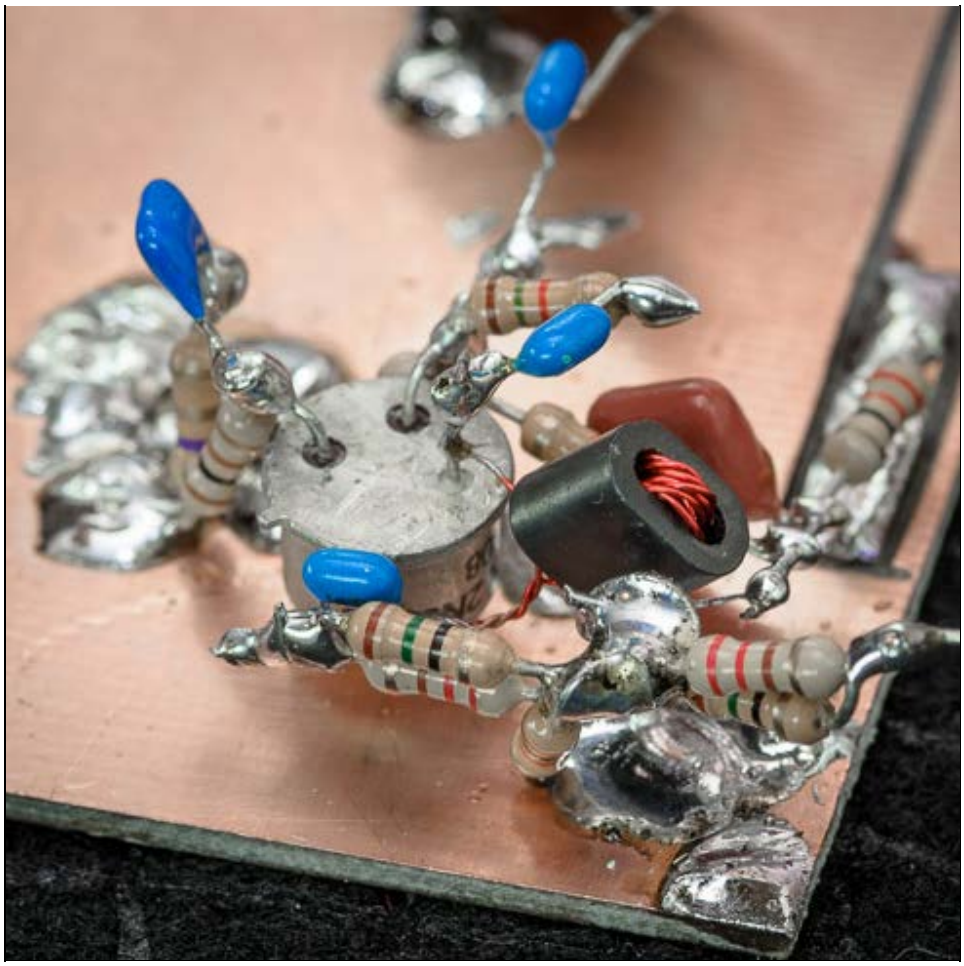
After peaking, I removed *all* the caps and measured them as shown in **red font**. I'll replicate these values with nearest standard-value fixed capacitors. The trimmer caps nearest to the ports proved more sensistive compared to the 2 middle trimmers capacitors during tuning.

This might be the coolest thing I've ever done on my test bench.

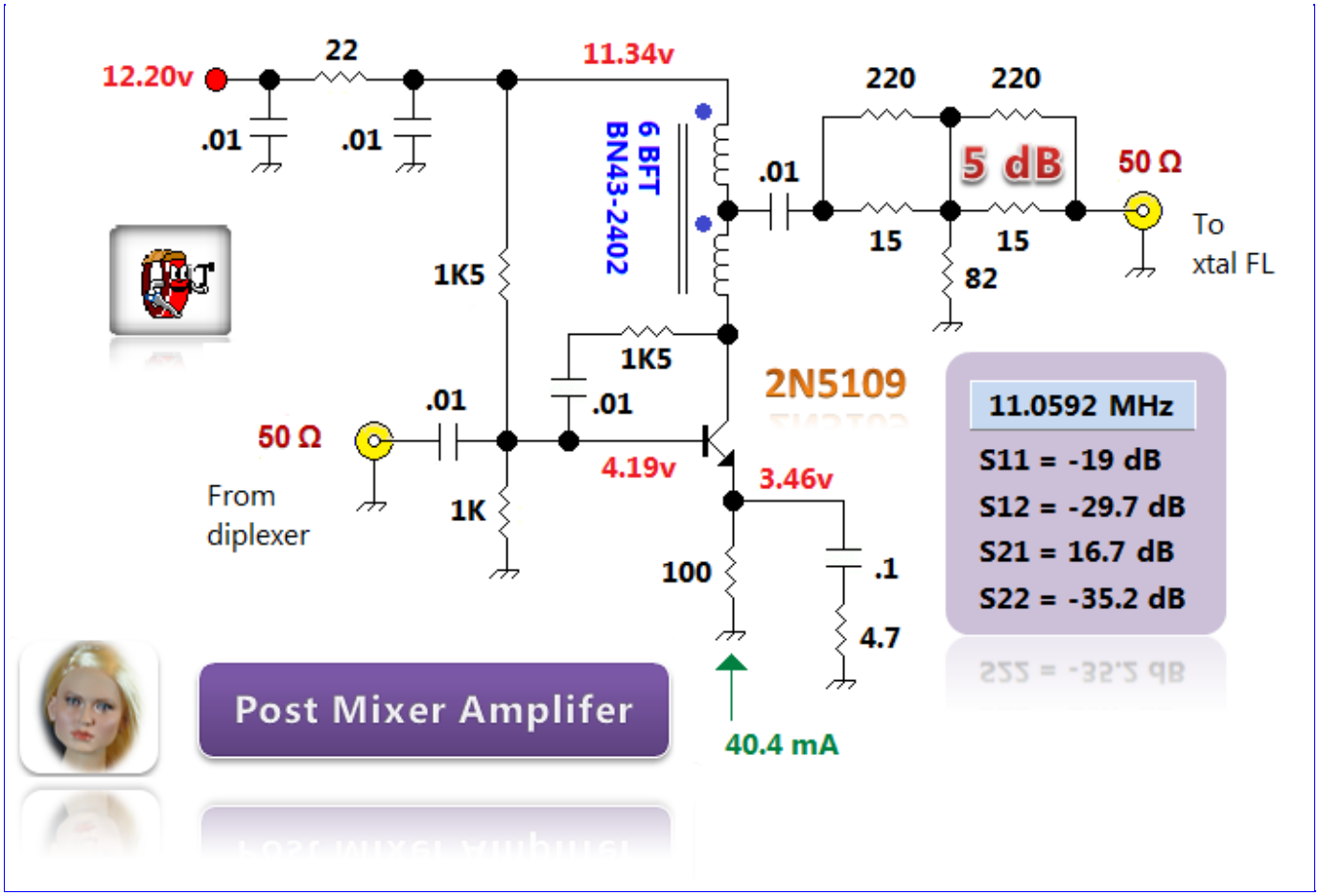


Above — My N = 3 filter breadboard. I left the capacitor leads long for re-use. I consume huge numbers of parts and recycle as possible to keep costs down.

Mixer and Post Mixer Amplifer Development



Above — Ugly breadboard shot with my ancient 105 mm lens for a change. It's a Beaverton Special feedback amp (FBA) biased for ~40 mA. For my post-mixer amp. If you need a stout RF amp, this one will do in a storm.

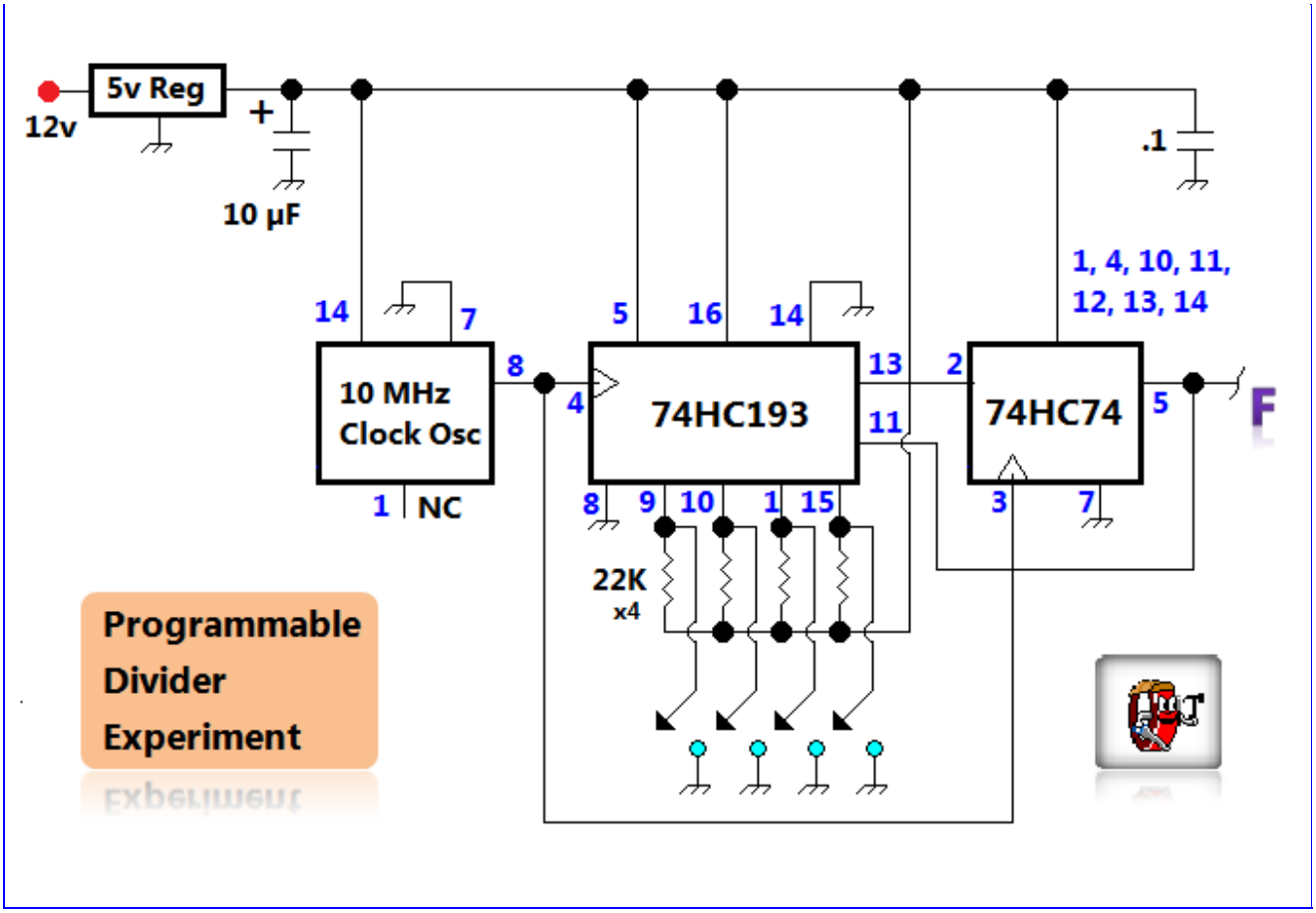


Above — The Beaverton Special feedback amp biased for ~40 mA with a 50 Ω output Z. The FBA with no 5 dB pad: S11 -18.5 dB; S12 -34.6 dB; S21 22.1 dB and S22 -31.8 dB. I love this FBA for its simplicity and well defined port impedance. My IF = 11.0592 MHz. Click for a discarded (but cool) photo.

To drive the 200 Ω input Z of my xtal filter I later converted the output to 200 Ω by changing the collector transformer and pad.

Frequency Synthesizers and Logic

I've spent a great deal of time reading about & playing with some HC series logic I bought in 2011-2012. I even purchased a few 74AC74, a D flip-flop you can clock up to 160 MHz or so. Most of my reading concerned frequency synthesizers: PLLs, dividers, prescalers, etc. Since most modern devices run a clock at GHz frequencies, the evolution of frequency synthesis technology from around the mid 1970's to now makes for a fascinating read.



Above — Using a CMOS 10 MHz clock oscillator, I tested out a positive edge-triggered programmable frequency divider designed by Wes, W7ZOI in 2011. It works perfectly.

Switches	Frequency (KHz)	N
0000	0	0
0001	3333	3
0010	2500	4
0011	2000	5
0100	1666	6
0101	1429	7
0110	1250	8
0111	1111	9
1000	1000	10
1001	909	11
1010	833	12
1011	769	13
1100	714	14
1101	666	15
1110	625	16
1111	588.2	17

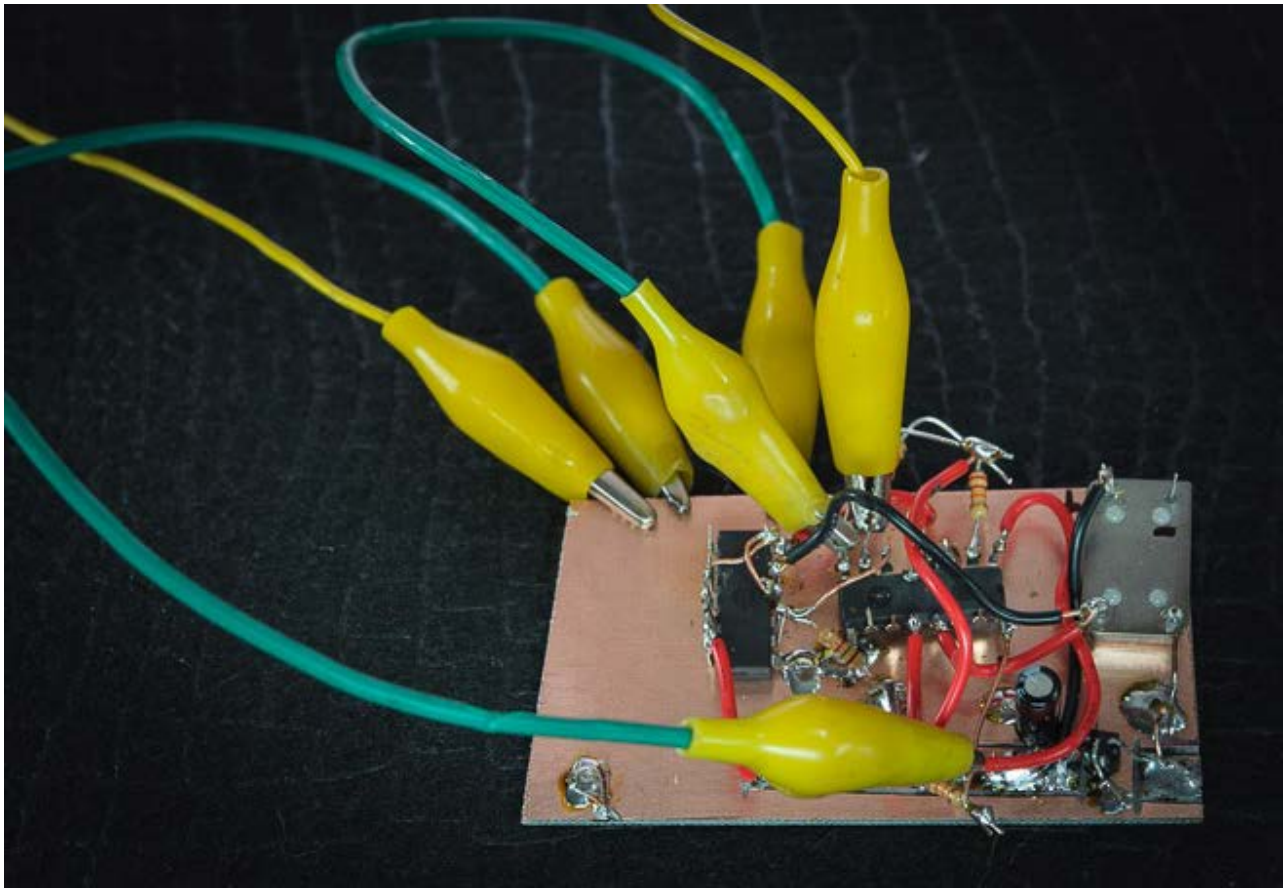
Programming Pin Sequence on 74HC193: 9, 10, 1, 15

Above — A test of all 16 combinations of the 4 programming switches with the 10,000 KHz input. I show the measured output frequency in KHz and the integer generated by dividing the output F into 10,000 KHz. You may leave pin 9 HIGH and just run it as a divide by 10 to 17 counter with 3 panel mounted programming switches — that's what I hope to do.

While working at Tektronixs, Wes designed the frequency synthesizer for the 492 Spectrum Analyzer; a portable SA that went to 21 GHz or so. His experiences informed the design of this simple, low-cost, edge-triggered divider.

Here, Wes applies the 74HC193 as a down counter where a number is loaded and then it count downs until zero is reached. The number gets boosted by 2 — 1 comes from the 74HC74 D flip-flop, while the other arises from the phasing of the signals. The 74HC74 makes the overall output coherent with its clock and avoids the flicker noise that otherwise might be generated in the long divider chain of the ripple counter.

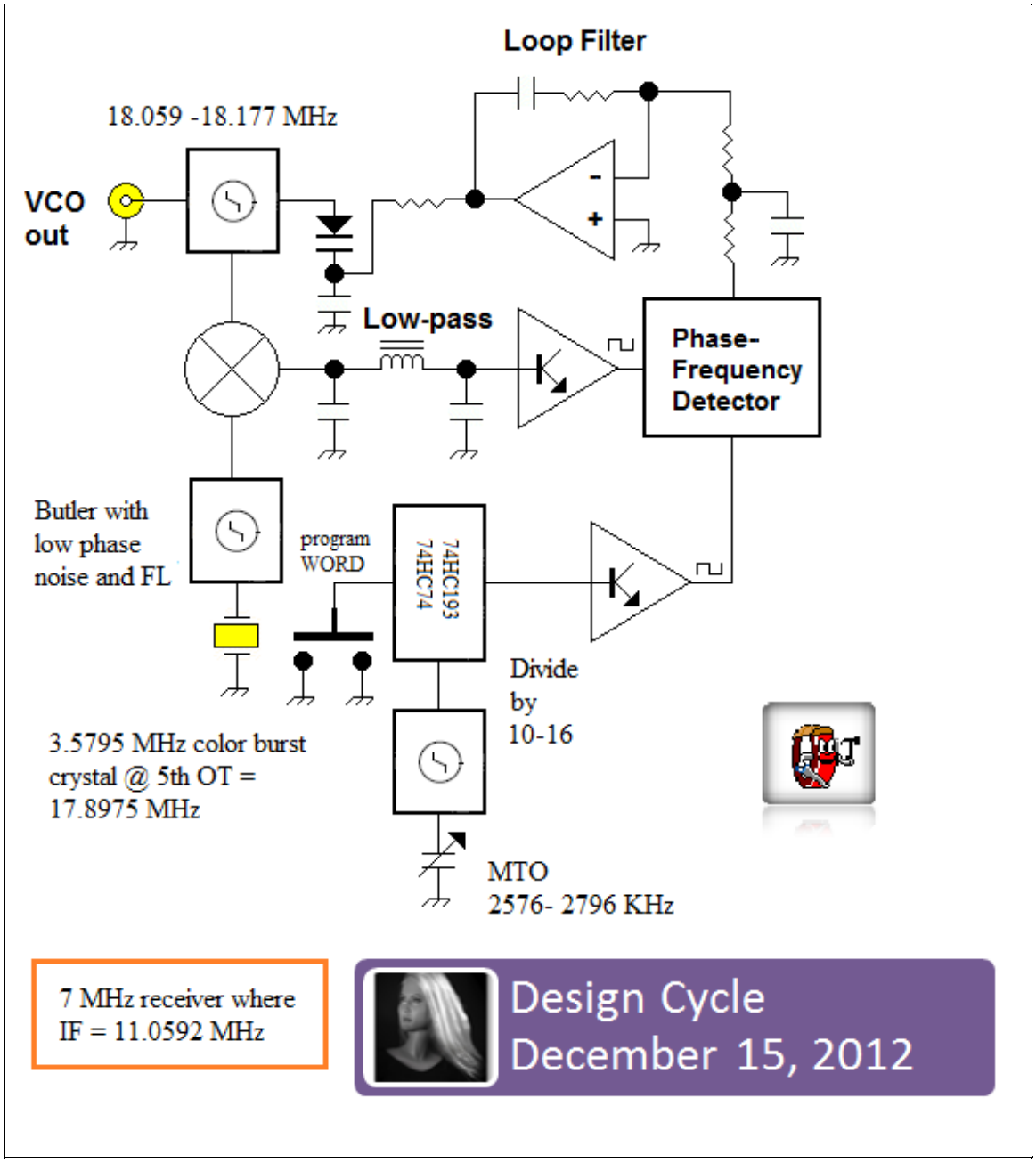
The standard binary to decimal conversion is altered in this circuit.



Above — My ugly breadboard. Instead of switches, I relied on bench jumpers to set a pin LOW.

Check out the standard binary switch combinations on this cool website: Math Is Fun. I also enjoyed their hexidecimal drum machine tool - Click. Actually, the whole web site looks great and I will recommend it to our friends with children and maybe even a few radio enthusiasts.

The Local Oscillator



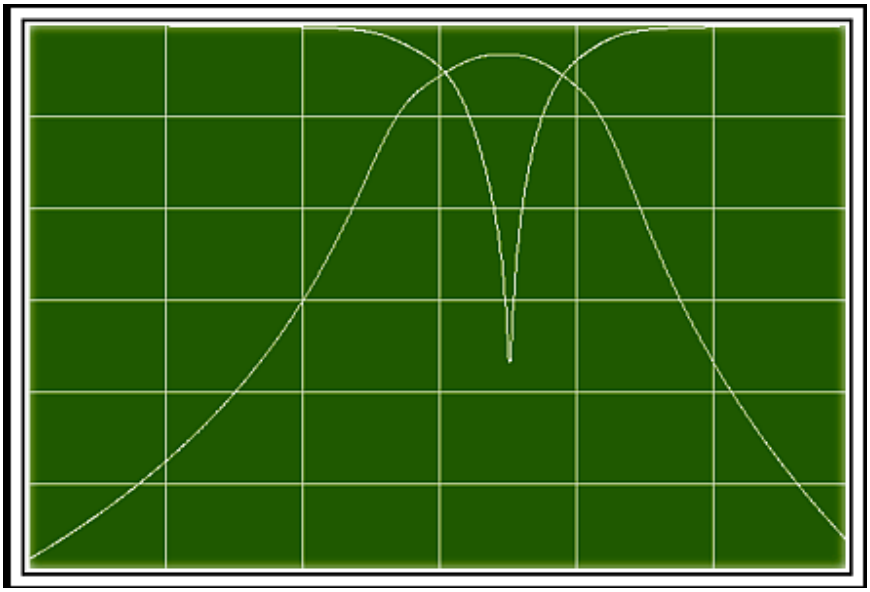
Above — An Almost Synthesizer based upon the design currently in use at W7ZOI.

I picked the MTO, xtal oscillator and VCO frequencies and will design the loop filter once I get the parts and a few small die-cast boxes on hand. The MTO division integer is low and quiet — plus this scheme enhances MTO frequency stability and spectral purity. A 4-bit ripple counter + flip flop perform the division where N = 10 - 16. I might go higher with N , but don't want to get too greedy. For CW, I'm generally in the bottom 20 KHz of the band anyway.

The MTO tunes about a 13-14 KHz range and the N gives 6 different tuning ranges (with some overlap) to span a total range of ~118 KHz. The 4-bit WORD is changed by 3 chassis-mount toggle switches and give a PLL reference of ~~175 to 280 KHz.

This almost synthesis seems a little crude in a time when people use microcontrollers to input a WORD into a chip with with the simple turn of a rotary controller, however this set-up fosters learning, huge fun, and when the oscillators plus loop filter are well designed; low phase noise and reciprocal mixing.

The Crystal Filter



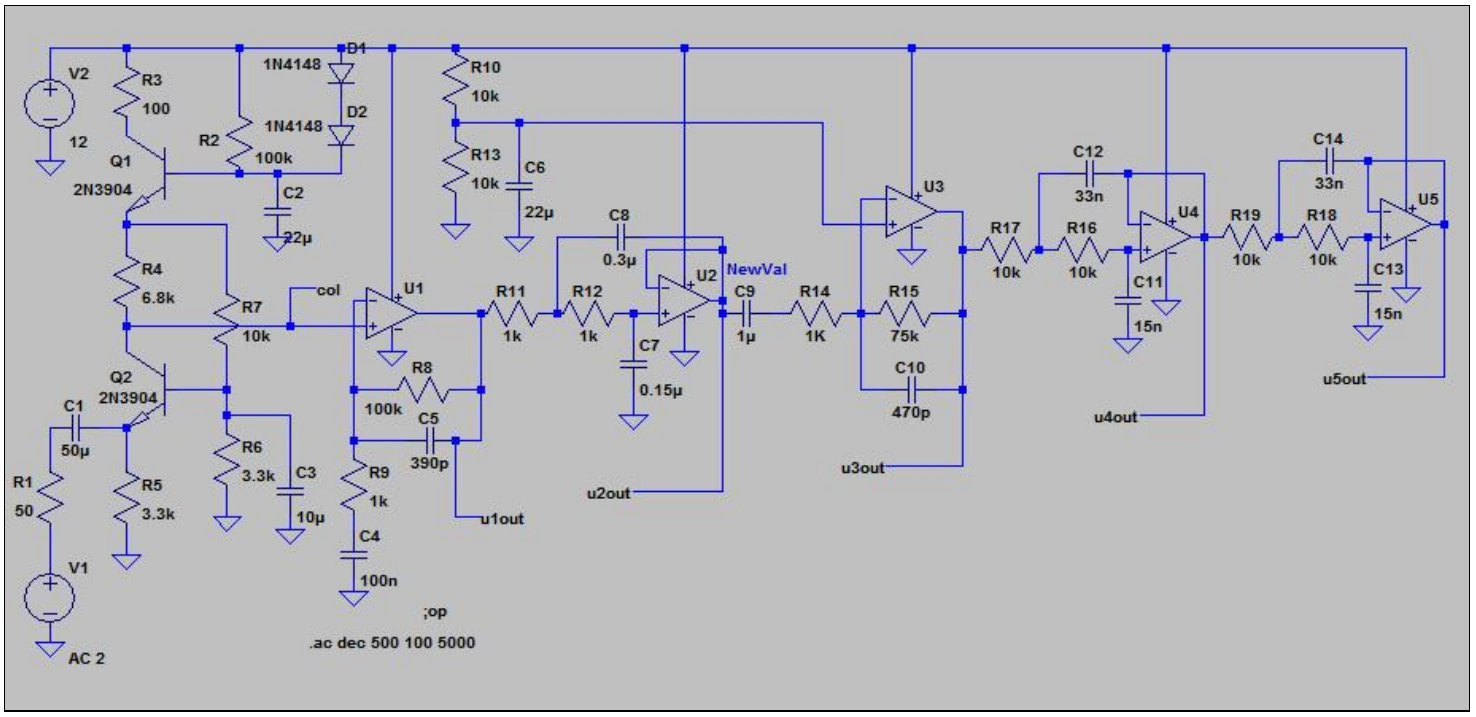
Above — A GPLA sweep of my 4 crystal filter where C parallel = 4.86 pF, Lm = .01 H, IF = 11.0572 MHz, BW = 1 KHz and R term = 200 ohms. This Gaussian to 6 dB filter design was chosen for speaker audio that sounds full + dynamic without the hollow ringing so often heard in homebrew receivers. I filter in my brain and first learned to copy CW across a room while listening to a speaker.

My code mentor Doug, could make coffee, walk around the room, talk to us and still 100% copy CW at 40 WPM. He taught us the ear to brain bandwidth is ~ 100 Hz when copying weak signal CW through a sideband crystal filter. To this day, I prefer listening through wider CW filters with brain band-pass filtering.

KL7R Memorial Supplemental Page

This web page is a supplement to the K7LR Memorial Receiver Experiments Web Page

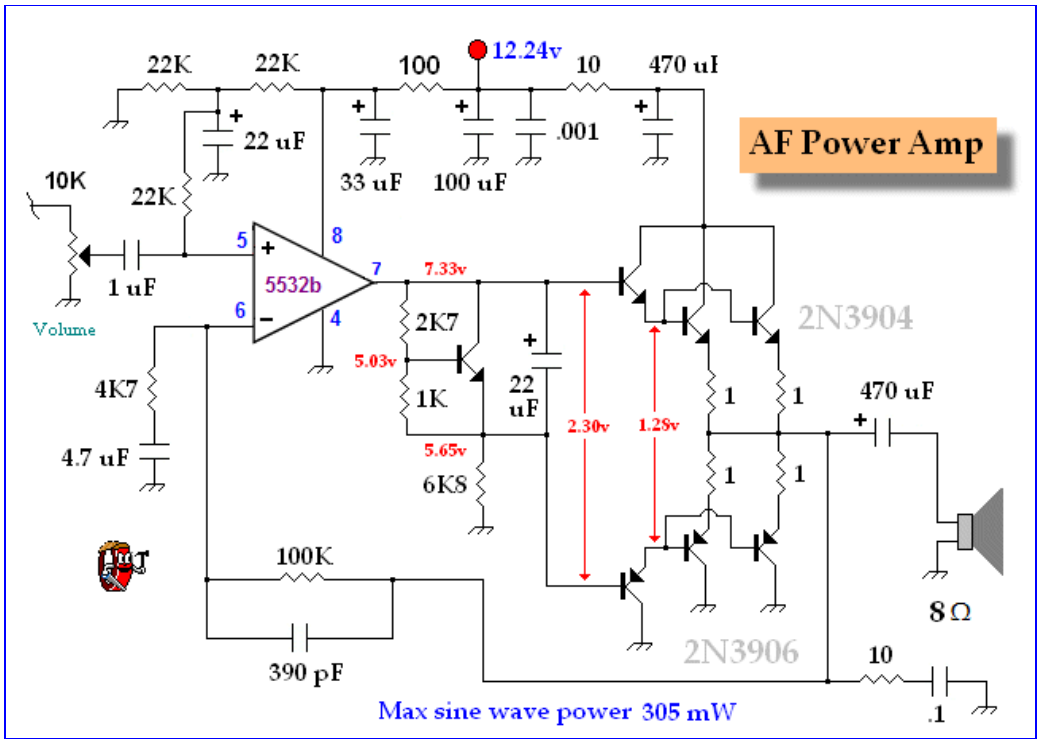
Audio Preamplifier Focus



Shown above is a SPICE build of an earlier prototype of the audio preamp chain which was ultimately used in this receiver. This SPICE work was performed by Wes, W7ZOI. Note that at this point, I had 100K plus 390 pF feedback in U1 and a 75K plus 470 pF feedback resistors in the op-amp amplifiers (really massive gain). I spent 4-5 days trying different audio stages before settling on the circuits shown.

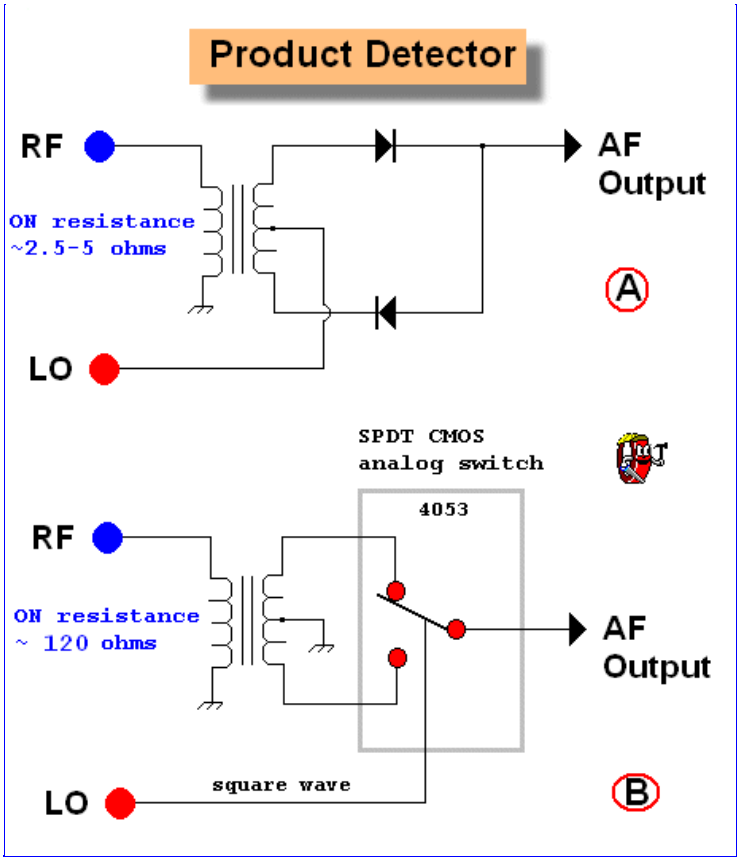
Shown above is the voltage versus frequency plot from a SPICE run of the schematic shown above. The gentle slope (decibels per octave) of the low pass filters can be seen. Certainly Chebychev filters with 1-2 dB of ripple would have provided steeper filtering, but the filters would have odd resistor values and be less easily modified or reproduced by popcorn builders. The V(u5out) shows a reasonable low pass response for the AF preamp.

Audio Power Amp

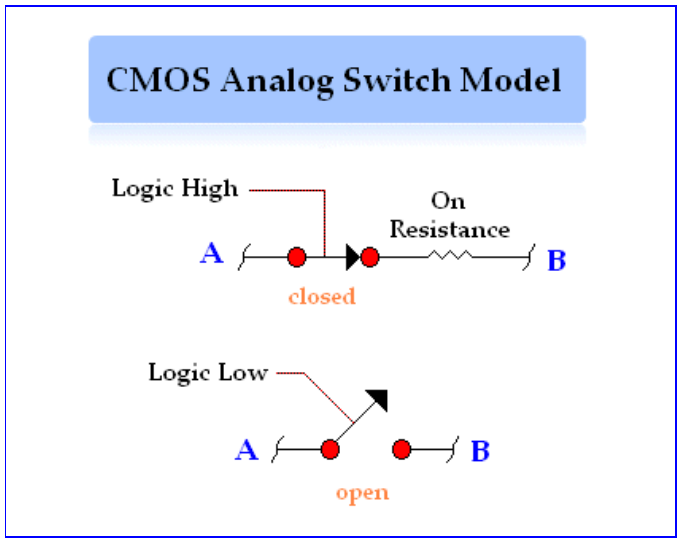


This schematic shows a stand-alone version of the audio power amp. If only 1/2 of the 5532 is used, connect pins 3 and 2 of the unused op-amp to the point where all the 22K resistors connect (VCC/2 bias point).

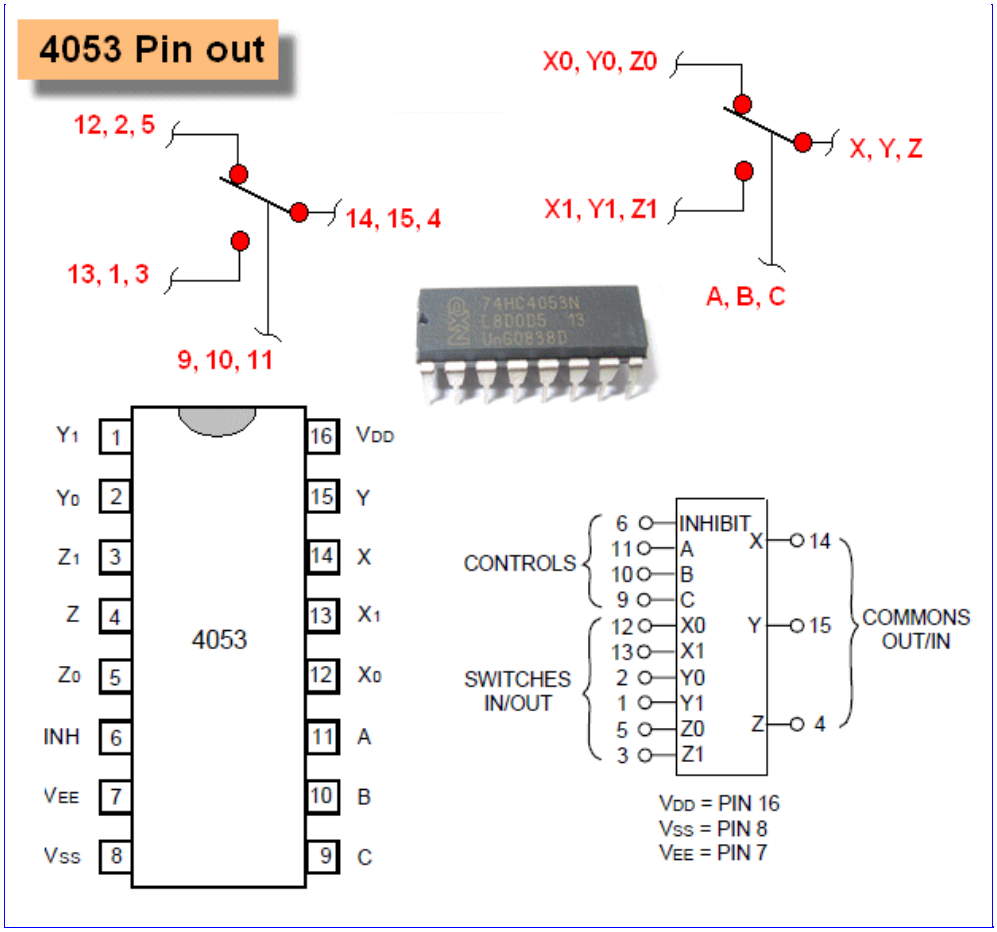
Product Detector Notes



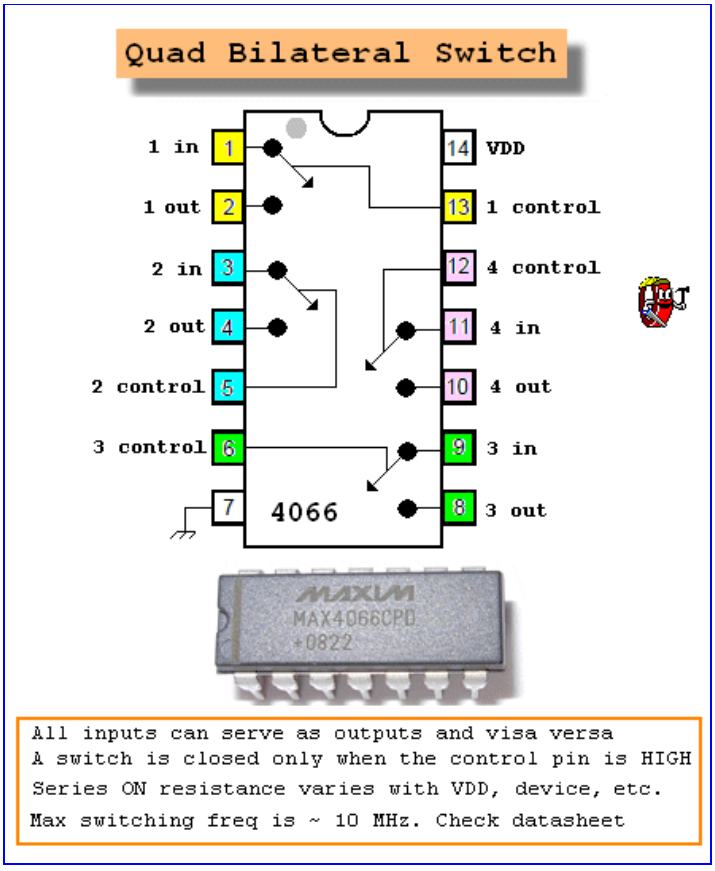
Shown above is the analogy between an SPDT CMOS switch and a single-ended, diode product detector. In both cases (A and B), the LO causes 1 diode (switch) to conduct, while the other is OFF. Then, on the other half cycle, the other diode (switch) conducts. If you wish to actually using the product detector/mixer depicted as "A" go with Figure 5.19B in EMRFD which has the LO applied to the transformer primary. As shown, "A" has poor LO to RF isolation, but it helps model the CMOS SPDT analog switch function. Function and performance are completely different topics... The on-resistance of the 4053 is much higher and thus performance is not identical.



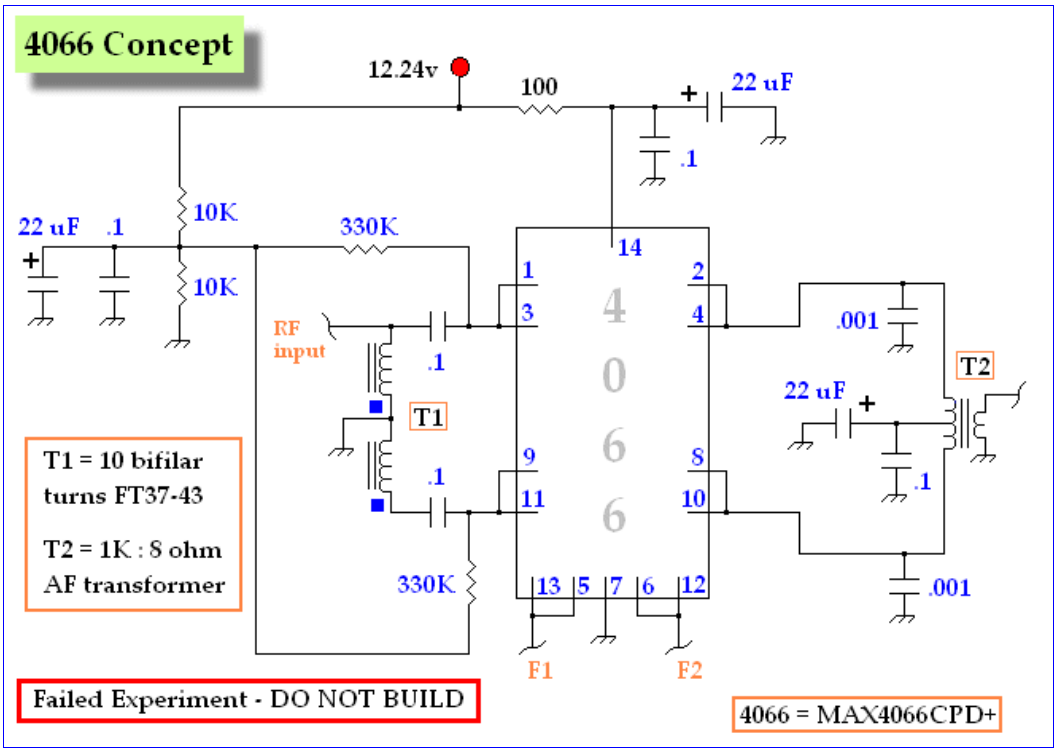
CMOS analog switch model. A or B are bidirectional and can be used interchangeably for the input or output. Ideally the on-resistance should be low to reduce propagation delay of digital signals plus resistance and perhaps distortion of analog signals. They can be used to switch digital bus data or analog signals. CMOS switches can replace mechanical reed switches in some low power circuits.



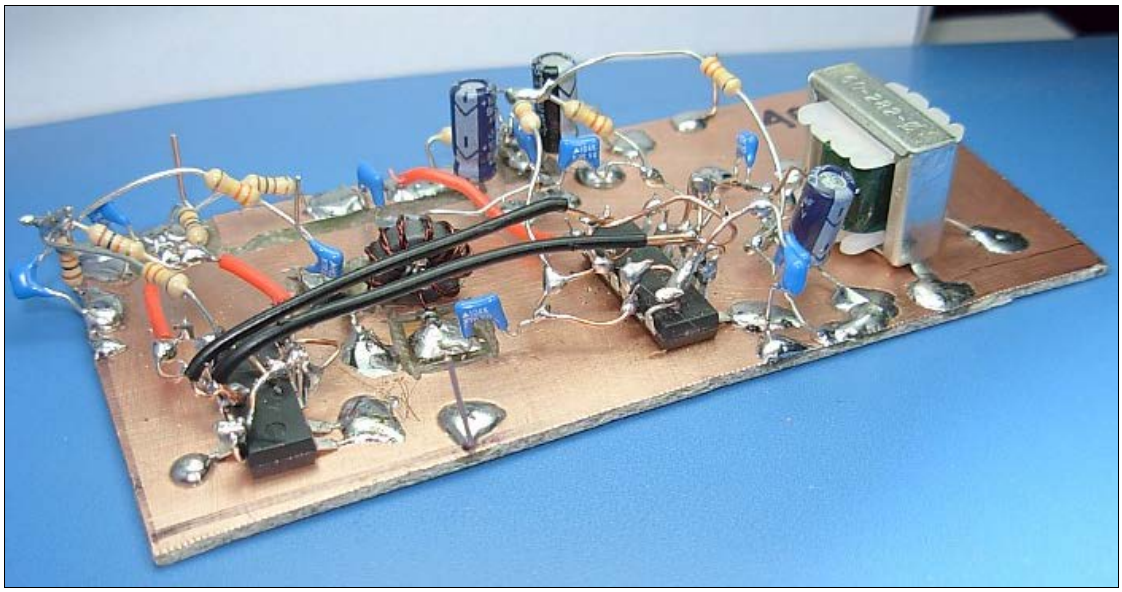
The CD4053BC is a triple 2-channel multiplexer having three separate digital control inputs; A, B, and C, and an inhibit input. Each control input selects one of a pair of channels which are connected in a single-pole double-throw configuration. A high on resistance makes the 4053 a compromise part for RF mixing. It might be perfectly okay for your design.



The 4066 is more easily understood when drawn in a semi-schematic. The on-resistance of this MAXIM version is better than the usual HC4066. About 45 ohms with a VCC of 12 or so.



A 4066 experiment that was a total failure. It is both humbling and useful to show your failures and not just the good fruit. This circuit had hum, noise and low output. The input transformer was a bifilar type, which was really cool.



The breadboard of the 4066 Concept schematic shown above.

The DIP IC switches used as product detectors in the K7LR memorial receiver experiments.

My work bench during some final experiments for this project.

K7LR MEMORIAL RECEIVER VFOs

This web page is a supplement to the K7LR Memorial Receiver Experiments Web Page

14 MHz VFO Circuit

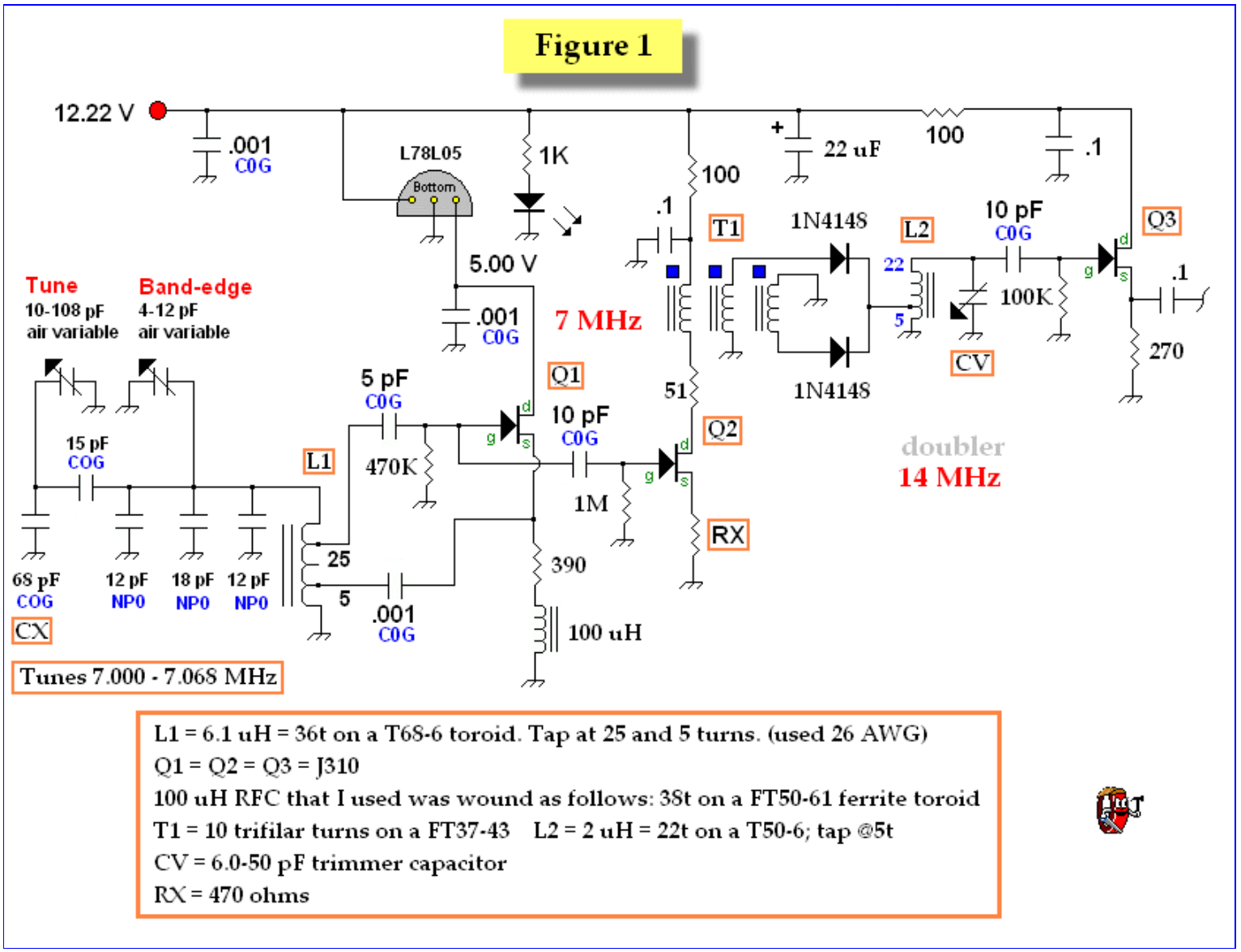


Figure 1 is the 14 MHz VFO used to switch (clock) the CMOS switch product detector. The tuning range can be easily changed by adjusting capacitor CX. As shown, the the tuning capacitor has a frequency range of about 68 Hz. If you decrease CX to 47 pF, the tuning range increases

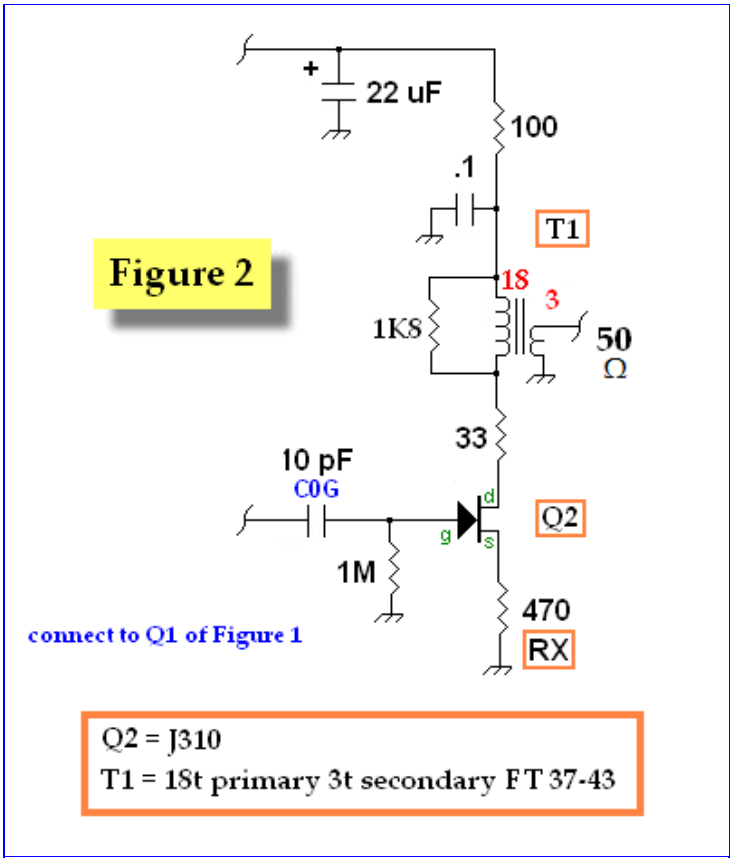
to around 99 Hz. Increasing CX to 100 pF gave about a 40 Hz tuning range. When you use a larger tuning range, fine tuning can become difficult. A geared reduction dial is one possible solution. For simplicity, I prefer to just keep the tuning range low as reasonably possible and use a big knob on the main tuning variable capacitor. Even the 68 Hertz tuning range shown can prove difficult for fine tuning. As experimenters, we are continually problem solving. Choose the tuning range and/or methods which suit your personal needs.

The base 7 MHz oscillator is doubled since a D flip flop is used to clock the CMOS switch product detector. This oscillator has low harmonic content and this is important for suppressing the 7 MHz fundamental frequency and its harmonics in the frequency doubler. Some astute builders may even match the two diodes used in the doubler. Diode matching is discussed on this page. CV tunes very sharply and the output at Q3 is a clean sine wave. This in turn provides a well balanced square wave at each output of the D flip-flop. The Q3 output voltage was measured at -0.46 dBm (0.6 volts peak to peak into 50 ohms).

Choosing Capacitor Values

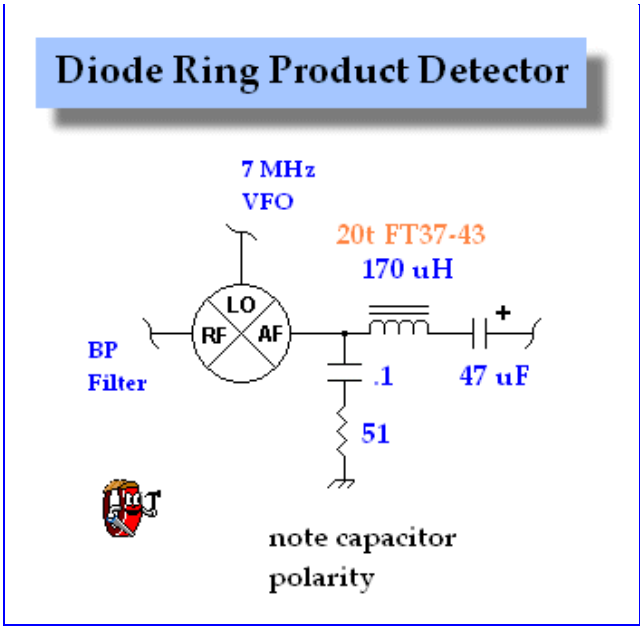
There are a number of ways to determine the capacitor values required to tune your VFO L-C tank circuit. It is pretty much essential to have a frequency counter and nice perhaps, to own a capacitance meter. Main tuning capacitors are typically harvested from an old radio or from the junk box. Its tuning range could be anywhere from 30 to 300 pF, or more. To limit the tuning range of this variable capacitor, normally you parallel a fixed-value capacitor and then series connect it to the top of L1 with a small, fixed-value capacitor. I performed this task entirely by trial an error using a frequency counter connected to the collector of Q2. Some fixed value NP0 or C0G capacitors plus a small trimmer capacitor are also required for tuning and to set the lower band edge respectively. Finding the right combination of capacitor values is painstaking, but with practice, gets easier. Once you have the basic capacitor values sorted out and your tuning range set, frequency stability experiments are then performed. This is known as VFO temperature compensation. Some times temperature compensation can be achieved by finding the right combination of NP0 and C0G capacitors. Additionally some negative or positive temperature coefficient capacitors may have to be soldered in and tested. Your final capacitor leads should be short as practically possible to reduce stray lead capacitance and for mechanical rigidity. Temperature compensation is discussed on this web page.

A 7 MHz VFO Circuit for Diode Ring Product Detectors



If you want to build a simpler version of the K7LR memorial receiver, a diode ring mixer may be substituted for the digital switch. The VFO requires modification as it will be run at 7 MHz. A different buffer is used and the Q2 to Q3 frequency doubler circuit is excluded. The diode ring product detector version is very nice. You could use a Mini-circuits TUF-1 , SBL-1 or alternate, or perhaps homebrew your own.

In Figure 2 is a VFO buffer configured to drive a 50 ohms input impedance, 7 dBm level, diode ring mixer. Change Rx to change the output voltage. With RX at 470 ohms, the output was somewhere around 5 dBm. This is useable for many situations. If you want exactly 7 dBm, the AC peak to peak voltage with a **51** ohm load resistor connected to the 3 turn link should be 1.43 volts. Adjust RX to achieve this voltage in your oscilloscope.



Note that if you build a diode ring product detector receiver version using the simple W7EL low pass termination network, the polarity of the 47 uF electrolytic capacitor will need to be changed as shown.

Key goals of this website include providing ideas, basic support and encouragement. I am delighted when builders make their own stuff and not copy my circuits. This is why we homebrew; to create, explore and share. Mike, K7LR did all of these very well.

Photography Links

Nikkor 50mm lenses

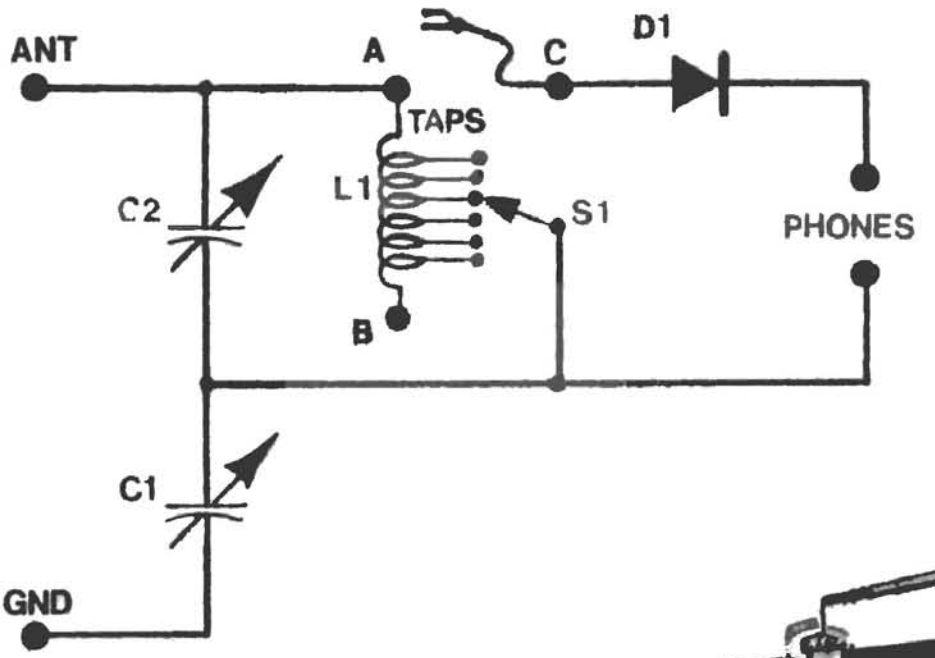
HENRI CARTIER-BRESSON - shot a lot with 50mm - video to watch over and over

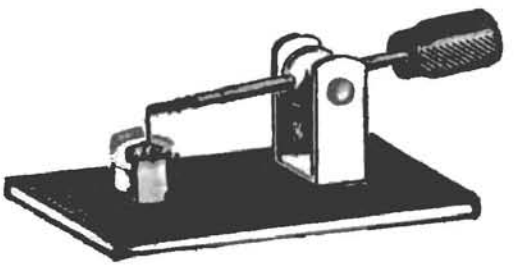
The genius of photography -- BBC4 series about the history of photography

http://www.freestylephoto.biz

all about CRYSTAL SETS

by
Charles Green





How-To-Build Simple Crystal Set Radios

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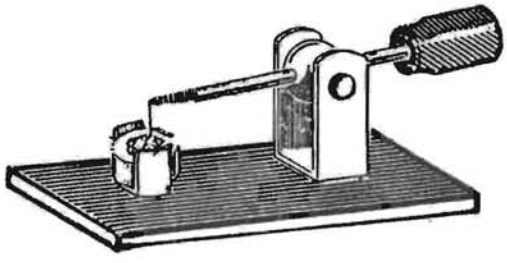
all about CRYSTAL SETS

by

Charles Green



How - To - Build Simple Crystal Set Radios

FIRST EDITION

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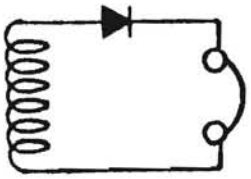
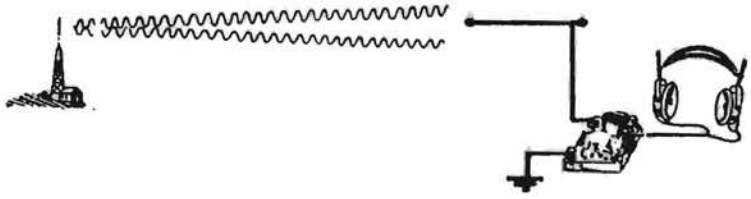


PREFACE

In the beginning of this century, when radio was called "Wireless", most of the early experimenters used a crystal set to receive radio signals.

The crystal set is still the simplest type of radio receiver. A catwhisker (made from a thin wire) touching a sensitive spot on a piece of galena or silicon mineral crystal and a tuning coil, form the basic parts of this radio.



Radio signals from a nearby station are picked up via antenna and ground connections to the crystal set. The galena or silicon crystal detects the signals and the resultant audio is heard with headphones. No battery or electrical power is required.

This book tells you about crystal sets and shows you how to build them.

Charles Green

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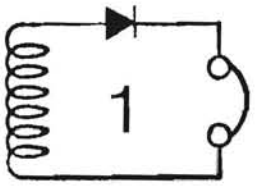
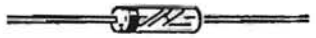
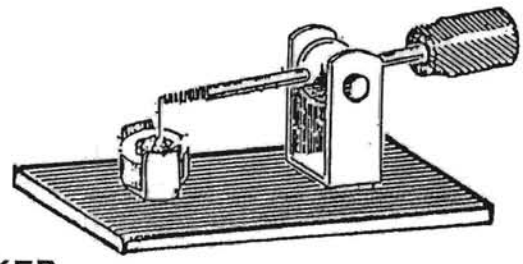


THE CRYSTAL DETECTOR

A crystal detector of radio waves can be either a traditional catwhisker and natural mineral crystal, or a modern semiconductor diode.



SEMICONDUCTOR DIODE



CATWHISKER DETECTOR

Certain natural minerals such as galena (lead sulphide) and silicon have electrically sensitive surface spots. In the beginning of this century, experimenters discovered that electrical current would flow into the sensitive spot on a mineral crystal from a catwhisker, but would not pass readily (high resistance) from the crystal to the catwhisker. This unique feature was used to detect (rectify) radio waves and extract the audio modulation.

HOW-IT-WORKS

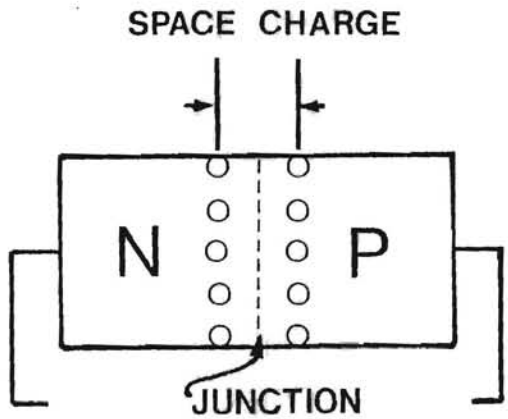
There are two general types of crystal detctors;

- Point contact diode
- Junction diode

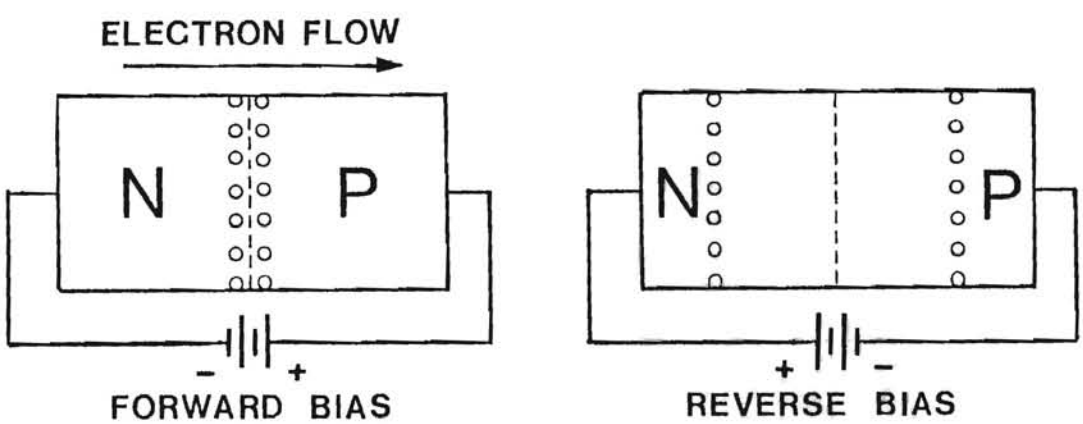
Both types use semiconductor material such as silicon, germanium, or galena. The basic semiconductor junction diode is made of two portions of the material "doped" in manufacturing by adding controlled amounts of chemicals. The doped portions have opposite electrical properties:

"P" TYPE, with an excess of positive electrical charges (holes).
"N" TYPE, with an excess of negative electrical charges (electrons).

Junction Diode



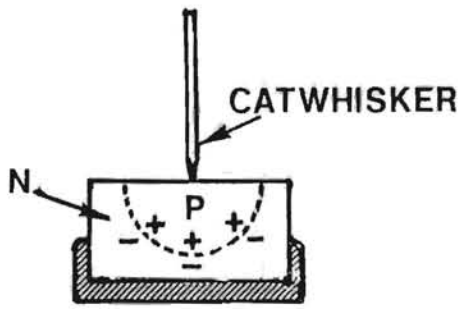
The boundary between the two doped portions is called a PN junction. The area on both sides of the PN junction is called the space charge region. When a positive voltage is applied to the "P" material (and negative voltage to the "N" material), the space charge region becomes narrow and electron current flows through the PN junction.



When the voltage polarities are reversed, negative to "P" and positive to "N", the space charge region widens (increasing the effective resistance of the PN junction) and the electron flow is very small (back current). If an ac voltage is applied to the PN junction, it will act as a rectifier (or radio wave detector).

Point Contact Diode

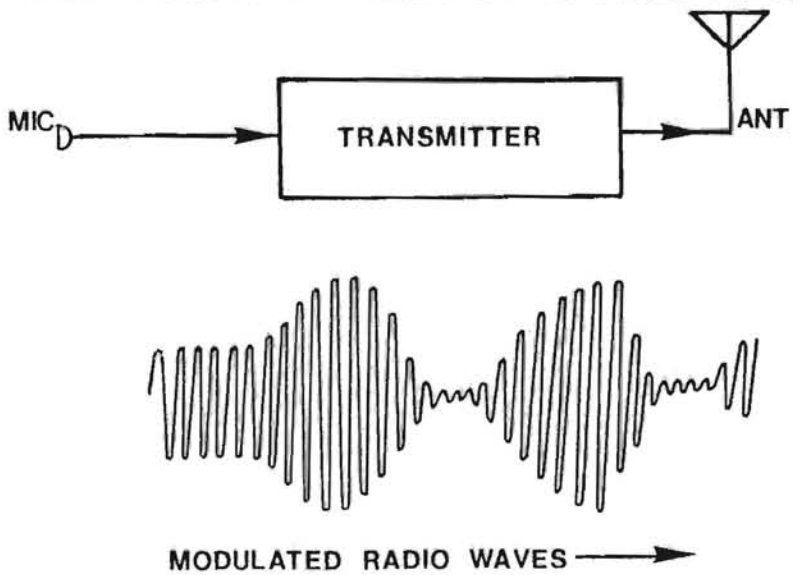
The point contact diode uses a "catwhisker" of thin wire contacting the surface of doped "N" semiconductor material. The pointed end of the wire produces a high intensity electric field in the area of the semiconductor surrounding the wire. When an electric current flows from the wire to the semiconductor material, a small region of "P" material forms around the area of the point contact. Thus a PN junction is formed which operates in the same way as the junction diode.



The sensitive spots of galena and silicon crystals used in traditional catwhisker detectors have naturally doped areas of semiconductor material, and operate in the same way as the point contact detector.

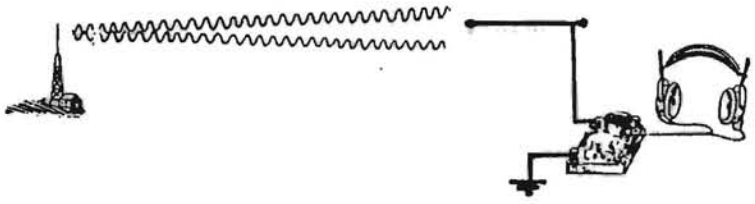
DETECTION OF RADIO WAVES

At the transmitting radio station, a microphone converts sound waves into electrical audio signals. The audio signals then modulate the transmitted radio waves.



the crystal detector

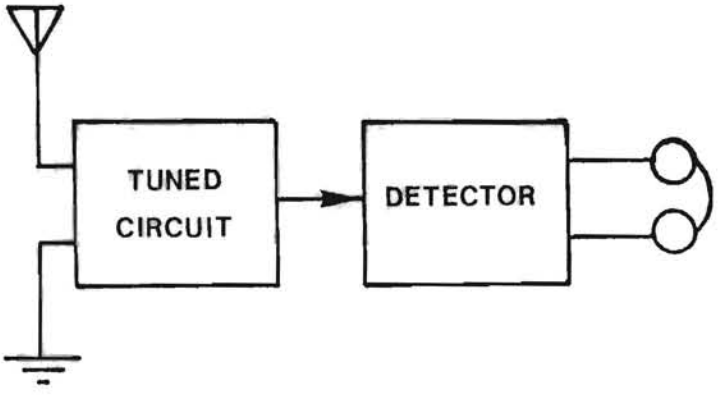
The transmitted radio waves are intercepted by the receiving antenna and small rf currents are induced in the antenna wire. The rf currents (high frequency ac) flow through the detector to ground. As previously explained, the crystal detector will pass current in only one direction, therefore, pulsating dc voltages (detected audio signals) are developed by the crystal equivalent to the audio modulation of the transmitted radio waves.



The detected audio signals flow through the headphone coils and affect the intensity of a magnetic field. This causes the steel discs of the headphones to vibrate and produce audible sound waves.

TUNING THE CRYSTAL SET

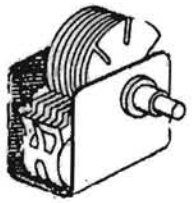
The antenna intercepts all radio waves in the area, and the signals will be heard jumbled up in the earphones. A nearby radio station may be received with the other stations heard in the background.



In order to receive a radio station properly, the crystal set must be tuned to the frequency of the transmitter. A tuned circuit composed of an rf coil (inductance) and a capacitor, is connected between the antenna/ground circuit and the detector.

Capacitor

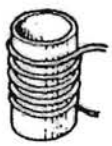
The modern capacitor is evolved from the old Leyden jar. It consists of two or more metal sections separated by a dielectric (air, paper, or plastic). The capacitor stores energy in the form of an electrostatic charge in its dielectric. A variable capacitor (used in a tuned circuit) has movable metal sections that adjust the amount of capacity.



A capacitor will allow pulsating dc current or ac current to flow through it, but will block steady dc current. A capacitor has an oppositon or "reactance" to ac that varies with frequency. The higher the frequency, the lower the reactance of a given value of capacity. Capacity is measured in micro-farads (uF) or micro-micro-farads (picofarads) (uuF or pF).

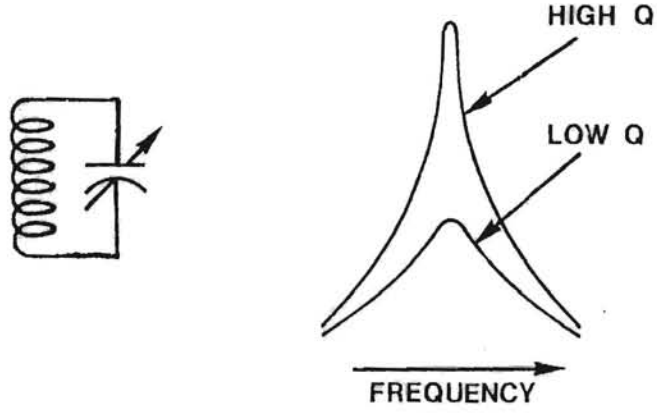
RF Coil

An rf coil is made of insulated copper wire wound on a plastic or cardboard tube. When electric current flows through the coil, a magnetic field is created. This magnetic field resists any change in electric current flow. This opposition or "reactance" to ac varies with frequency. The higher the frequency, the higher the reactance of a given value of rf coil inductance. Inductance is measured in millihenrys (mH) and microhenrys (uH).

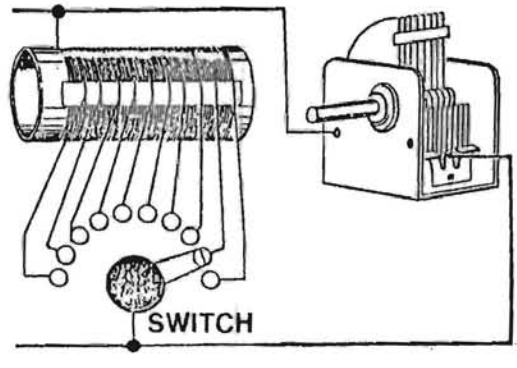


Tuned Circuit

A tuned circuit is composed of a capacitor and an rf coil. As previously described, the reactances of a capacitor and an rf coil change with frequency (but in opposite ways). At a particular frequency, the two reactances will cancel each other out. This condition is called "resonance". The impedance of a parallel tuned circuit is at its highest at resonance. Radio signals at the resonant frequency will be greatly magnified, while signals at other frequencies will be shunted to ground (rejected).

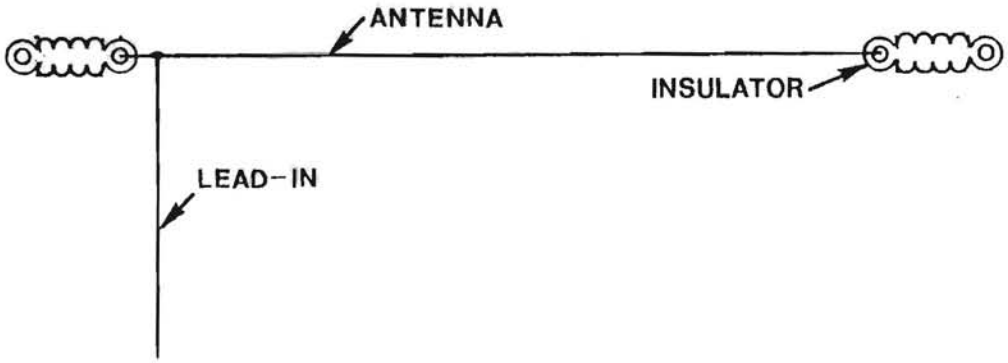


The resonant frequency of a tuned circuit can be changed by adjusting the capacitance (using a variable capacitor),or by varying the inductance (using a slider or taps on the rf coil). A ferrite core can also be used to change the inductance of a coil. The selectivity of a tuned circuit is dependent upon the quality (Q) of the components and the inductance /capacity ratio (L/C ratio). A high-Q tuned circuit is more selective than a low-Q circuit. A tuned circuit with a high L/C ratio (large inductance and small capacity) is more selective than a low L/C ratio circuit.



ANTENNA AND GROUND

A simple crystal set depends upon the strength of the radio waves received by the antenna to produce audio in the earphones. If you live in a strong signal area (a nearby radio transmitter), a crystal set can produce loud audio in earphones with only an inside antenna. But in lower signal areas, a good outside antenna with a ground connection is required.



If you are fortunate enough to be able to put up an outside antenna, the basic rule of thumb is the higher, the better reception. Check your local electronics store for an antenna kit. You can construct an antenna using a glass or plastic insulator at each end of a length of copper wire (insulated or bare). Make the antenna as long as possible. It does not have to be in a straight line to work.

CAUTION

ANTENNA INSTALLATION
NEAR ELECTRICAL POWER
LINES IS DANGEROUS.

Make sure that your antenna is far enough away from any power line, so that it will not touch a wire if it falls.

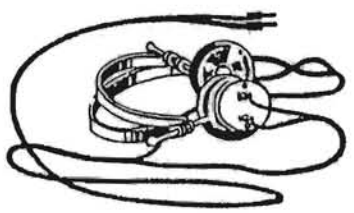
Connect a length of insulated wire between the end of the antenna and your crystal set antenna terminal. This antenna lead-in wire can be fed into your house through a window.

If you can not erect an outside antenna, run a length of insulated wire as long and high as possible through your house or apartment. It will work, but not as well as an outside antenna because of the shielding affect of the house walls.

Connect a ground lead to a cold water pipe. Make sure that it is a metal pipe. Or, you can buy a ground rod from your local electrical supply store and drive it into the ground near your house. Then connect your crystal set ground lead to the rod. For safety install a lightening arrestor between the antenna lead and ground (see the instructions with the device).

HEADPHONES

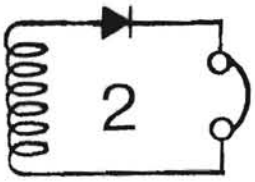
Headphones (also called earphones, or headset) are required to hear the received audio from a crystal set. The standard headphones are made with dual coils of fine wire connected in series in each phone. Horseshoe magnets are built into each phone case to provide a magnetic field. A disk of thin sheet iron (diaphragm) is mounted near the top of the coils. The phones are connected in series with a set of flexible leads terminated in metal "phone tips".



Detected audio signals flow through the coils and affect the intensity of the magnetic field from the horseshoe magnet. This changes the magnetic attraction of the diaphragm, and causes it to vibrate and produce audible sounds.

The total resistance of the series headphone coils is used as a measure of audio sensitivity. A resistance of 2000 ohms is sufficient. More sensitive headphones have a resistance of over 5000 ohms and are used to hear very weak signals.

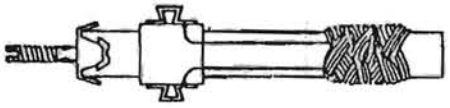
The 2000 ohm headphones can be obtained at electronic parts mail order houses.



SIMPLE RECEIVER PROJECTS

SIMPLE CRYSTAL SET

Look at the schematic diagram. The crystal set uses a tunable ferrite coil L1, and a crystal diode D1. This type of tuning coil is called a "loopstick". The ferrite core is screwed into the coil form and increases the inductance of L1, thus tuning the receiver to a lower frequency.



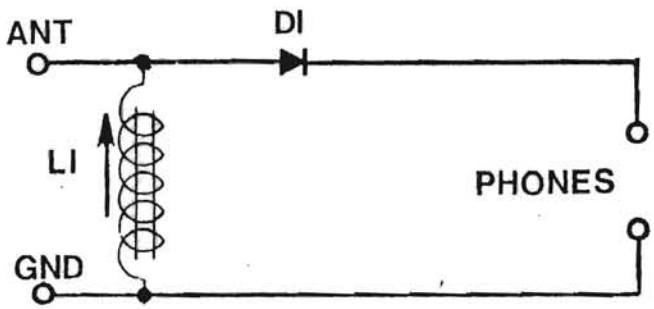
CONSTRUCTION

As shown in the illustration, the crystal set components are mounted on a 5 1/2-inch by 2 1/2-inch by 3/4-inch wood base. Layout and install the crystal set parts in the locations shown. The loopstick L1, is mounted with a metal strip supplied with it. Push the snap-in end of L1 into the large hole of the metal strip. Then bend 1/2-inch of the other end of the strip to form a bracket and mount it on the base with a wood screw. If the metal strip is not available, make one out of sheet metal approximately 2-inches long by 1/2-inch wide. Then drill a hole to fit L1 at one end, and a small hole to fit a wood screw at the other end.

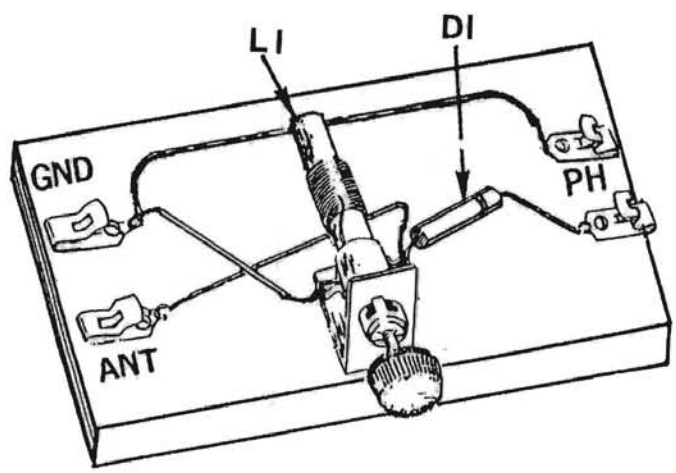
Mount the four fahnestock clips and solder lugs with wood screws in the locations shown in the illustration.

WIRING

Use hookup wire to connect the crystal set components as shown in the schematic and illustration. The loopstick L1 may have three terminals. One of the terminals is a tap, which is not used in this circuit. Make sure that you connect only to the terminals at the start and end of the loopstick coil.



SIMPLE CRYSTAL SET



List of Materials

D1 - Germanium diode
(1N34A or equiv.)
L1 - Loopstick
(see text)
4 - Fahnestock clips
BASE - 5 1/2 X 2 1/2 X
3/4-inch wood section
PHONES - 2000-ohm headphones
MISC. - Solder lugs, wood screws,
knob and bracket to fit L1,
hook-up wire.

Install a small knob on the end of the L1 adjustment screw. Make sure that the screw turns freely with the knob. Position the 1N34A diode as shown before soldering the leads.

Experiment 1

Connect the antenna, ground, and 2000-ohm headphones to the crystal set terminals. Rotate the L1 knob counter-clockwise as far as it will go. Do not turn past this point, or you may break the ferrite rod inside the coil form. The ferrite is now outside of the coil winding. This is the highest frequency point of the tuning range of L1.

Slowly turn the L1 knob clockwise and listen for received radio stations. When the L1 knob is turned fully clockwise, the ferrite will be completly inside the coil (increasing the inductance) and it will be at the lowest frequency point of the tuning range.

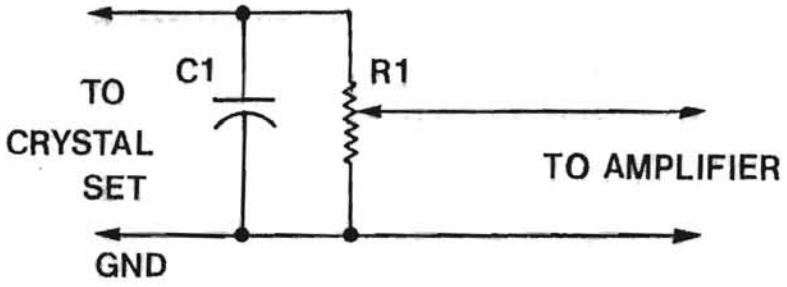
Experiment 2

Connect capacitors, with values from 50 mmf to 300 mmf, in parallel with L1. Then tune L1 and note how the frequency range has changed.

AMPLIFIER CONNECTION

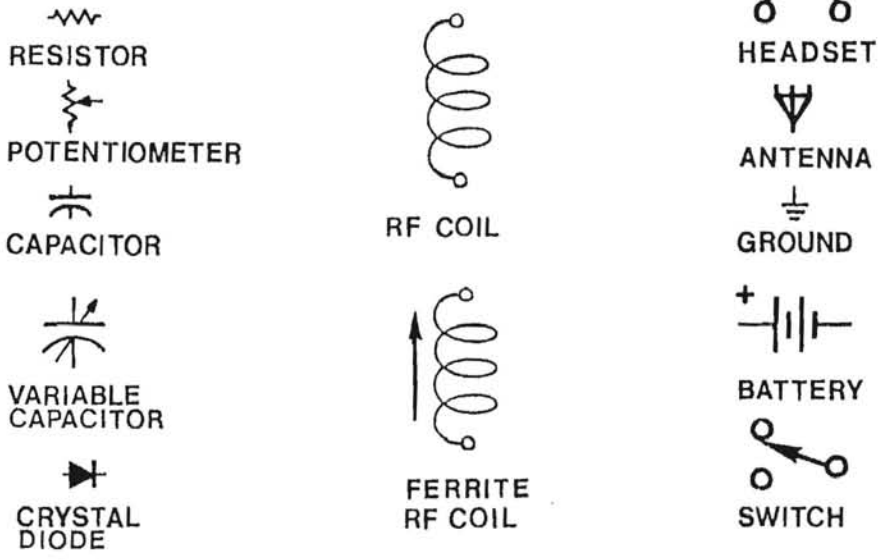
The simple Crystal Set (and the other receiver projects) can be connected to an amplifier in place of earphones. This amplifier can be a single channel input of your home stereo.

Parts values are not critical; C1 is approximately .005 mfd, and the volume control R1 can be any value from 1000 ohms to 10,000 ohms. Use shielded wire between R1 and the amplifier input jack.



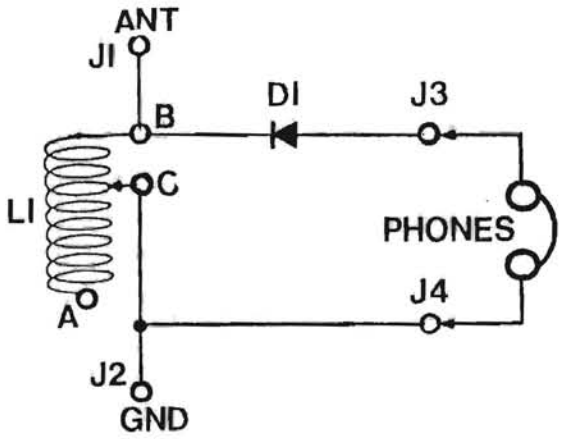
Amplifier Connection

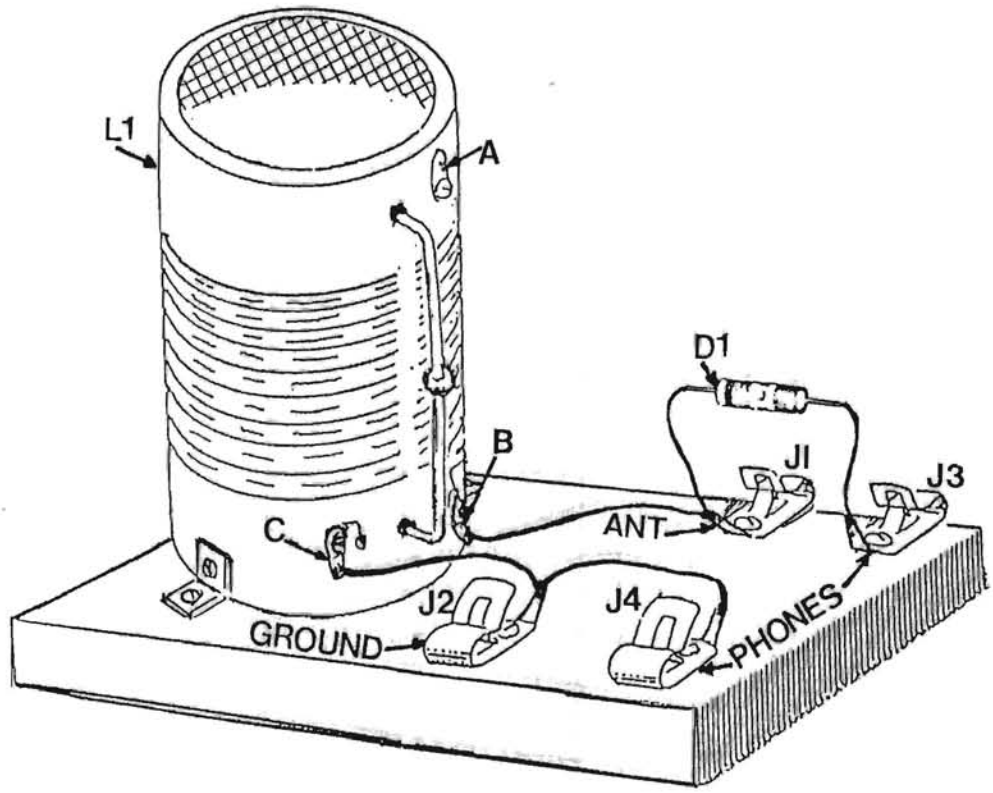
Schematic Symbols



List of Materials

D1 - Germanium diode, type 1N34A, 1N60, 1N82A, or equiv.
L1 - Slider-tuned coil (see text)
4 - Fahnestock clips (terminals)
BASE - 6 x 3 x 1/2-inch section of wood (see text)
PHONES - 2000 ohms headphones
MISC. - Coil form (see text), No. 24 enameled copper wire, 3/32 inch dia. brass rod, brass ball slider (see text), solder lugs, two small brackets, hook-up wire, wood screws, machine screws and nuts.

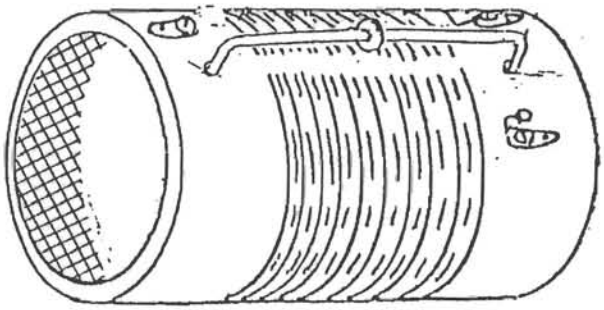




SLIDER TUNED CRYSTAL SET

SLIDER TUNED CRYSTAL SET

In the early days of radio, slider tuned coils were used in crystal sets. This method of tuning the inductance uses an adjustable metal contact (slider) that moves over the coil. The slider makes contact with individual coil wires to tune in the radio stations.

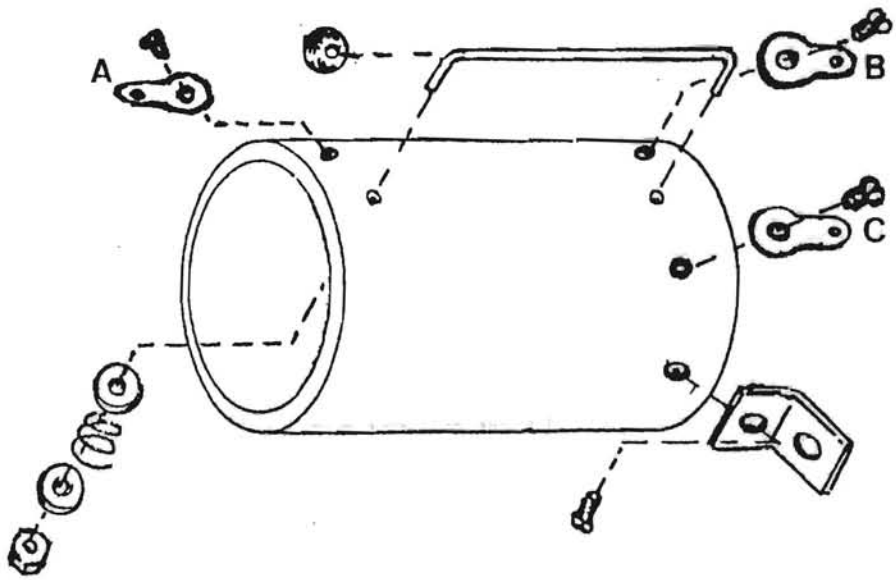


The enamel insulation of the coil wires in the slider path is removed for good electrical contact.

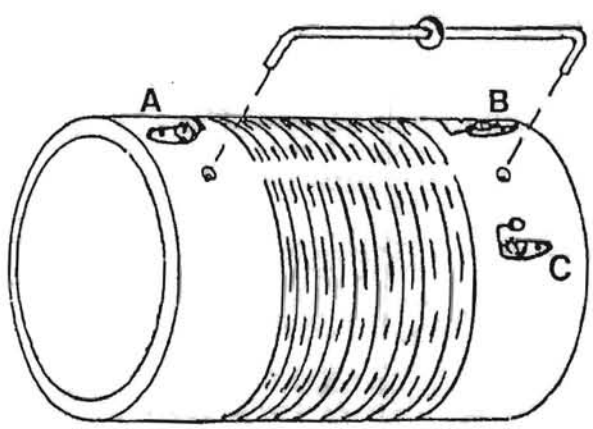
You can experiment with a slider tuned coil by building this crystal set. As shown in the illustration, the coil is mounted upright on a 6-inch by 3 1/2-inch by 1/2-inch wood base. A semiconductor diode is used for a detector.

TUNING COIL CONSTRUCTION

Look at the drawing of the tuning coil L1 construction details on the next page. The tuning coil is wound on a cardboard or plastic mailing tube section, 2 5/8-inches in diameter and 5-inches long. Make sure that you select a rigid, thick-walled tube for a coil form. The tube used in our model has a 3/16-inch wall. The coil form must be rigid to hold the slider ball and rod assembly.

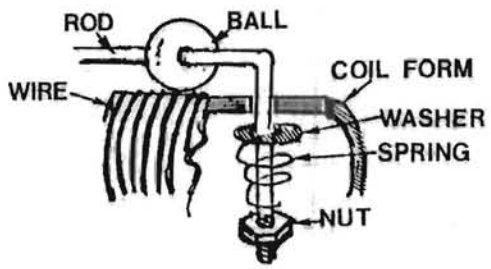


Before winding the coil, drill all the holes in the form as shown in the drawing, and then install the solder lugs. Position the solder lug ends toward the coil form edges.



The coil is wound with No. 24 enameled copper wire. Remove the enamel from the wire end with sandpaper and then solder the wire to the lug. Start winding the coil, keeping the turns as tight and close together as possible. Wind the wire to the front of the bottom solder lug B. Remove the enamel and solder the wire end to the lug.

A brass ball approximately 3/8-inches in diameter is used as the moving contact slider. This brass ball can be found at your local electrical supply store. It is a type of ball that is used as a top nut for small table lamps. Drill a hole through the ball so that it will slide over a length of 3/32-inch diameter brass rod. Bend the brass rod at each end (after placing the ball in the center of the rod), to fit through the holes in the coil form as shown in the drawing. Cut the rod 1-inch down from the bend at each end.



Insert the brass rod with the slider ball into the coil form holes. Then place a small washer and spring on one end of the rod (inside the coil form). Solder a nut on the end of the rod to hold the spring in place. Do the same to the other end of the rod. After soldering, slide the ball over the wires and check that it slides smoothly. Use sandpaper to remove the enamel insulation of the wires in the path of the ball slider. Connect a wire between the soldered nut on the bottom end of the brass rod and the solder lug near the rod.

BASEBOARD

All of the crystal set components are mounted on a wood baseboard as shown in the drawings. Cut the baseboard to size; a 6-inch by 3 1/2-inch section of 1/2-inch thick wood (maple is used in our model). Sand the edges after cutting. It is not necessary to paint or stain the baseboard.

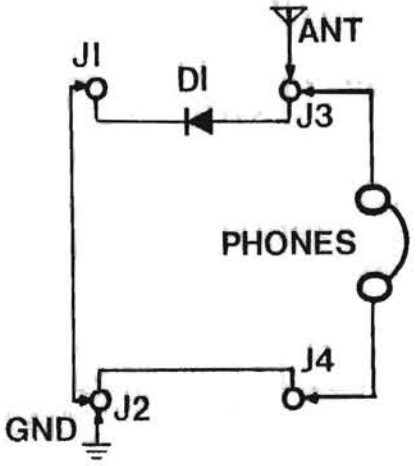
Mount the four terminals and solder lugs as shown in the drawing. Label the four terminals with either rub-on lettering or cemented paper tags. Install two brackets on opposite sides of the bottom of the coil, and mount it on the baseboard with the lugs toward the terminals.

WIRING

Bare No.22 wire was used to wire our model, but any kind of hookup wire can be used. Connect the wires between the solder lugs as shown in the drawing, except do not connect the antenna terminal lug wire to the coil terminal at this time. Connect the crystal diode between the solder lugs as shown.

Experiment 1

This experiment will show how an untuned crystal detector functions.



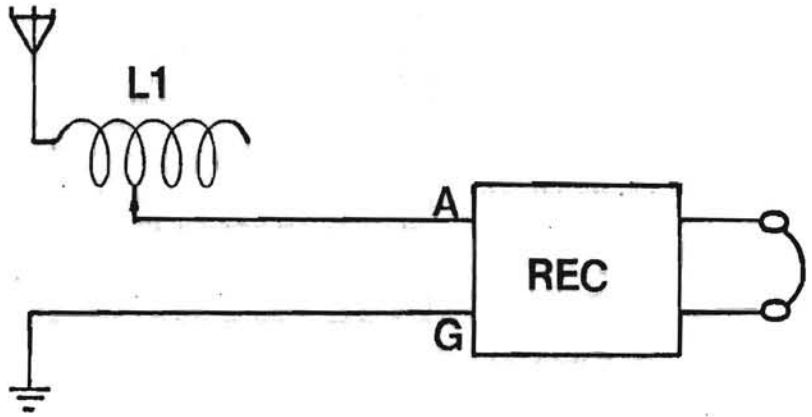
Connect the untuned detector circuit as shown in the drawing, and then hook-up your antenna and ground leads. Connect 2000-ohm headphones to the circuit and listen to the detected radio waves.

You will probably hear a strong local radio station, with other, weaker radio station transmissions all jumbled together.

Experiment 2

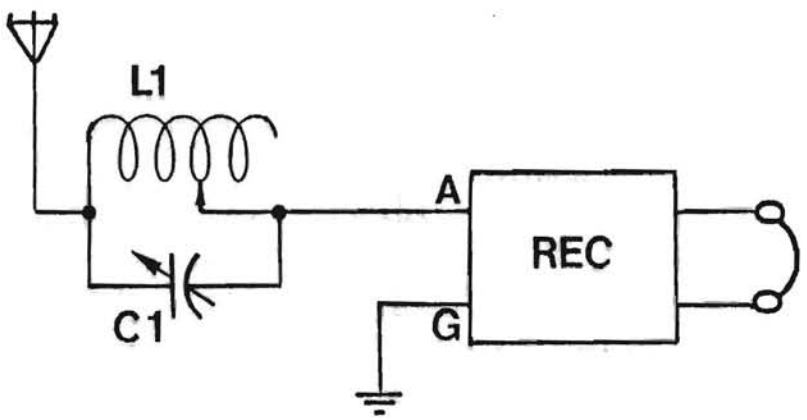
Reconnect the baseboard wires to the circuit of the Slider-Tuned Crystal Set as shown in the drawing. Listen to the detected radio waves as you tune the coil by sliding the brass ball over the sanded wire path. You should be able to receive the radio stations much better than the untuned detector circuit of Experiment 1.

LOADING COIL

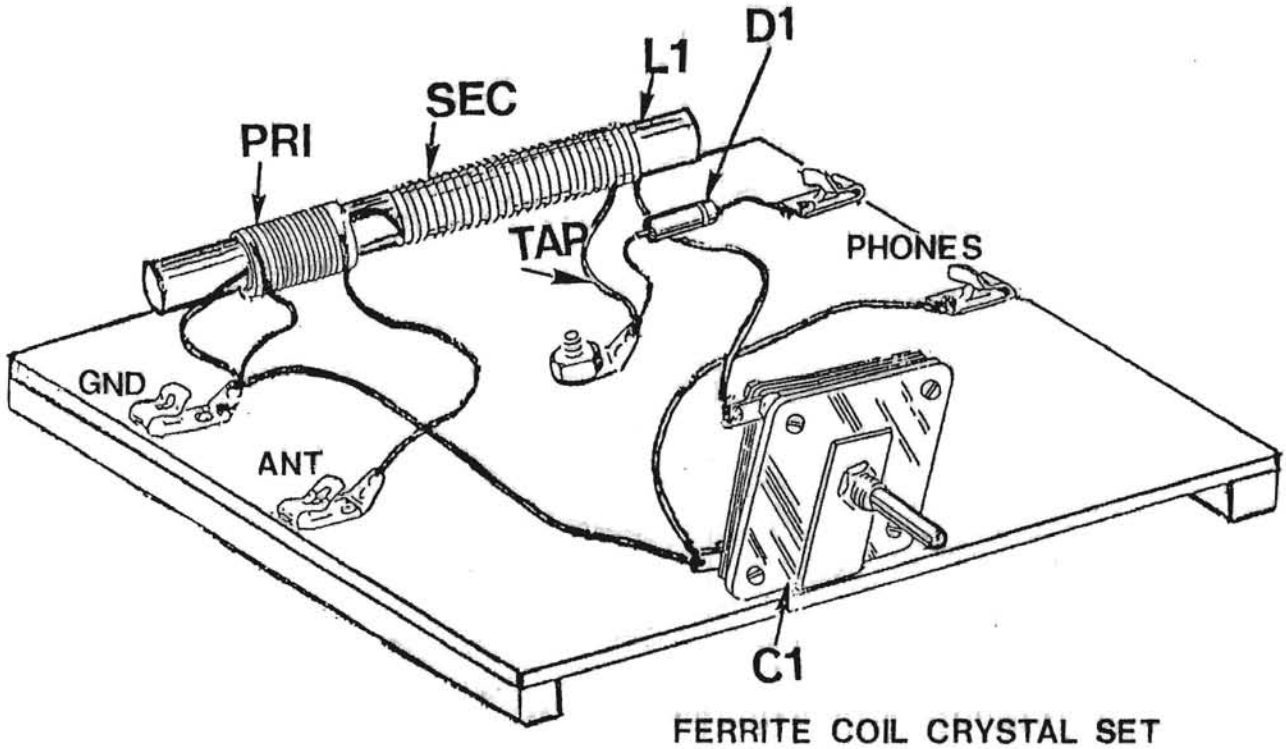


An additional slider tuned coil, L1, can be used as a Loading Coil to increase the electrical length of your antenna. This may allow better reception of the lower broadcast band frequencies. Connect L1 in series with the antenna lead to the receiver, and tune it for best reception.

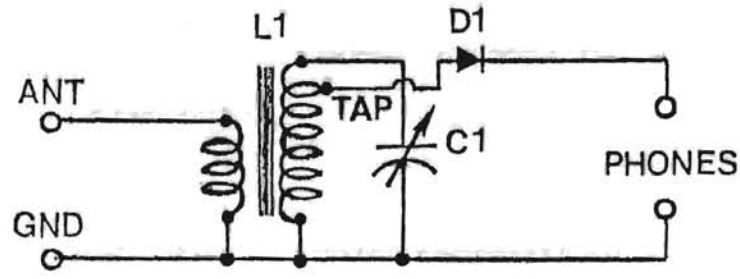
WAVE TRAP



L1 can also be used in a Wave Trap circuit to tune out a strong local signal, and allow easier tuning of weaker stations. C1 is a 365 mmf variable capacitor connected in parallel with L1. Adjust L1 and C1 so that the interfering station is at minimum.



FERRITE COIL CRYSTAL SET

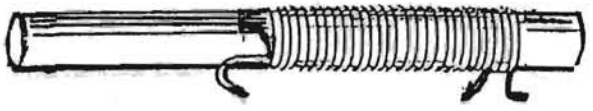


List of Materials

C1 - 365 mmf variable capacitor.
D1 - Germanium diode (1N34A or equiv.).
L1 - Ferrite coil (see text).
4 - Fahnestock clips.
2 - wood sections; 6 X 1/2-inch square (see text).
BASE - 8 1/2 X 6-inch sheet plastic (see text).
PHONES - 2000-ohm headphones
MISC. - plastic tube section for L1 primary winding (see text), solder lugs, machine screws and nuts, wood screws, bracket for C1 (see text), knob for C1, hookup wire.

FERRITE COIL CRYSTAL SET

This crystal set has a large ferrite core coil L1 of a type that is normally used for a built-in antenna in portable radios. This type of coil has a very high "Q" and may be more selective than an air-core coil. As you can see in the schematic, a primary coil is wound around the ferrite coil. The original winding is used as the secondary.



The primary coil (connected to the antenna and ground) can be moved close to, or away from, the secondary (original winding connected to D1). This variable coupling is used to control the selectivity of the crystal set.

PRIMARY COIL

The primary coil is wound on a 1-inch section of plastic tube that can fit loosely over the ferrite rod. A 1/2-inch diameter plastic tube is used on the receiver shown in the illustration. Wind 12-turns of hookup wire evenly over the tube and twist the ends together to prevent movement. Leave 4-inch leads from the primary coil winding. Slide the primary coil over the center of L1.

BASEBOARD

The crystal set components are mounted on a 8 1/2-inch by 6-inch section of sheet plastic. The thickness of the plastic is not critical, but it must be rigid enough to hold the parts without bending. A 6-inch long 1/2-inch square wood section is installed under each end of the baseboard. A section of wood can be used for the baseboard with wood screws to mount the parts, in place of machine screws and nuts.

The model shown uses a plastic variable tuning capacitor, but an air type capacitor can be used for C1 instead. Make a metal bracket approximately 1-inch wide by 1 1/4-inch high and a 1/2-inch right angle mounting foot from sheet aluminum. Drill a hole in the top of the bracket to fit C1 and two holes in the mounting foot to fit small machine screws. Locate the components on the baseboard as shown in the illustration and then drill mounting holes. If you have a cardboard or plastic mounting supplied with ferrite coil L1, you can mount it on the baseboard. If not, L1 can be taped loosely on top of the baseboard. Th ferrite coil L1 in our crystal set was obtained from an old radio. Install the components on the board with solder lugs, machine screws, and nuts.

WIRING

Connect the parts with hookup wire as shown in the illustration. Use the schematic diagram as a guide. Make sure that the ground lead from L1 secondary passes inside the primary coil. Also check that the primary coil slides freely over the L1 winding. Install a knob on the C1 shaft, and mark the terminal designations on the baseboard.

Experiment 1

Hookup the antenna, ground and 2000-ohm headphones to the crystal set terminals. Slide the primary coil over the center of the L1 winding. Tune C1 for radio stations. Then slide the primary coil towards the ground end of L1 while tuning C1. Note the difference in selectivity.

Experiment 2

Change the number of turns of the primary winding and repeat Experiment 1.

Operating Tips

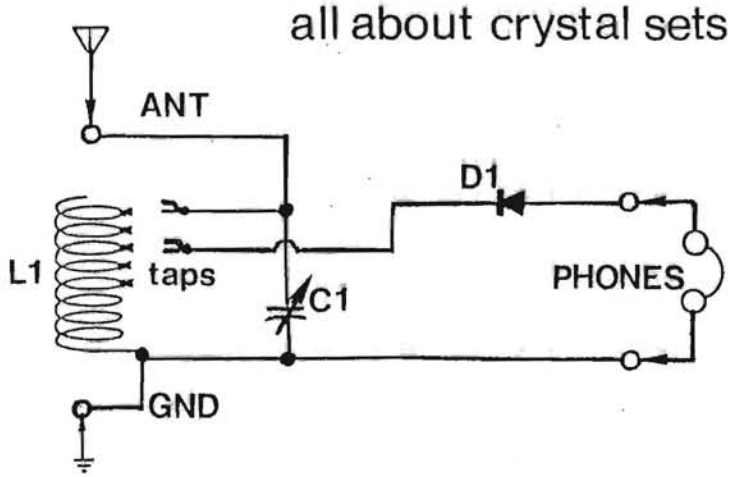
After operating a crystal set for some time, you may notice a difference between day, evening, and night reception of radio stations. This is due to the two ways that radio signals travel between the transmitting antenna and a receiving antenna.

(a) The ground wave; which moves along the surface of the earth.
(b) The sky wave; which travels upward and is reflected by the ionosphere.

The reception of broadcast stations is dependent upon the time of day. During daylight hours, stations are received primarily by ground wave transmission. At night, the broadcast stations will be received via sky wave transmission. Also, some radio stations change to a lower power at night. Reception in the evening may be more difficult as the fading ground wave is interfered with by reflected waves. But later, it may be possible to hear far-away stations transmitted by sky wave.

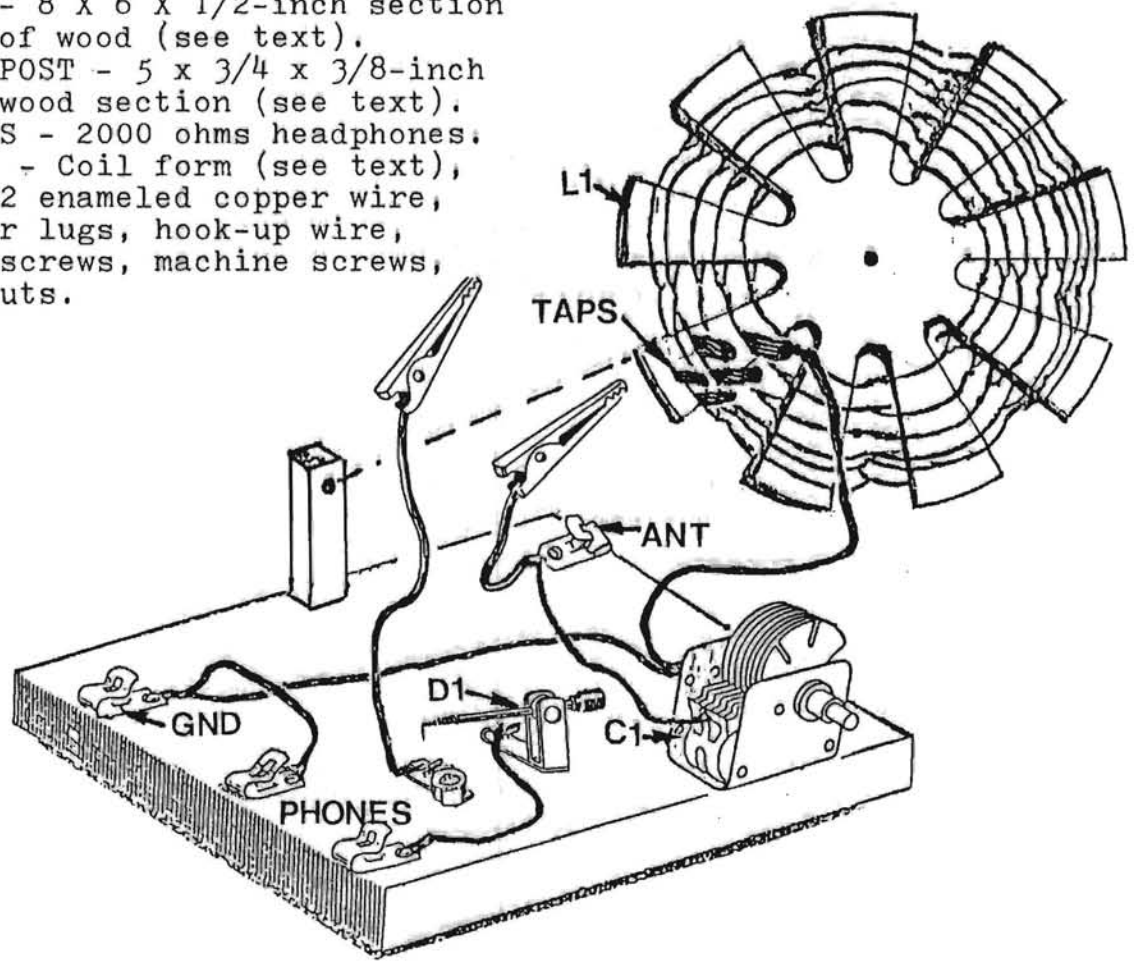
Your geographical location to radio stations is very important for reception, as a crystal set receiver does not amplifiy signals. It is entirely dependent upon the strength of the radio waves. If you are in a deep valley surrounded by high mountains, reception will be difficult. If you live in a flat plain, with no obstacles to radio transmission, reception may be very good.

An outside antenna is necessary for for best reception. If possible, try erecting an extra antenna at a right angle to the original one, and switch it in to provide a directional effect for increased signal strength. Keeping a log of received stations for a week may be helpful in determining the radio reception conditions in your area.



List of Materials

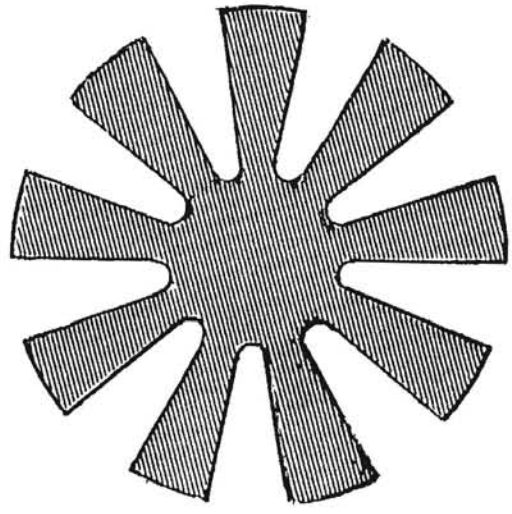
C1 - 365 mmf variable capacitor.
D1 - Catwhisker crystal detector (or 1N34A or equiv. diode).
L1 - Spiderweb coil (see text).
4 - Fahnestock clips.
2 - small alligator clips
BASE - 8 X 6 X 1/2-inch section of wood (see text).
MTG. POST - 5 x 3/4 x 3/8-inch wood section (see text).
PHONES - 2000 ohms headphones.
MISC. - Coil form (see text), No. 22 enameled copper wire, solder lugs, hook-up wire, wood screws, machine screws, and nuts.



SPIDERWEB COIL CRYSTAL SET

SPIDERWEB COIL CRYSTAL SET

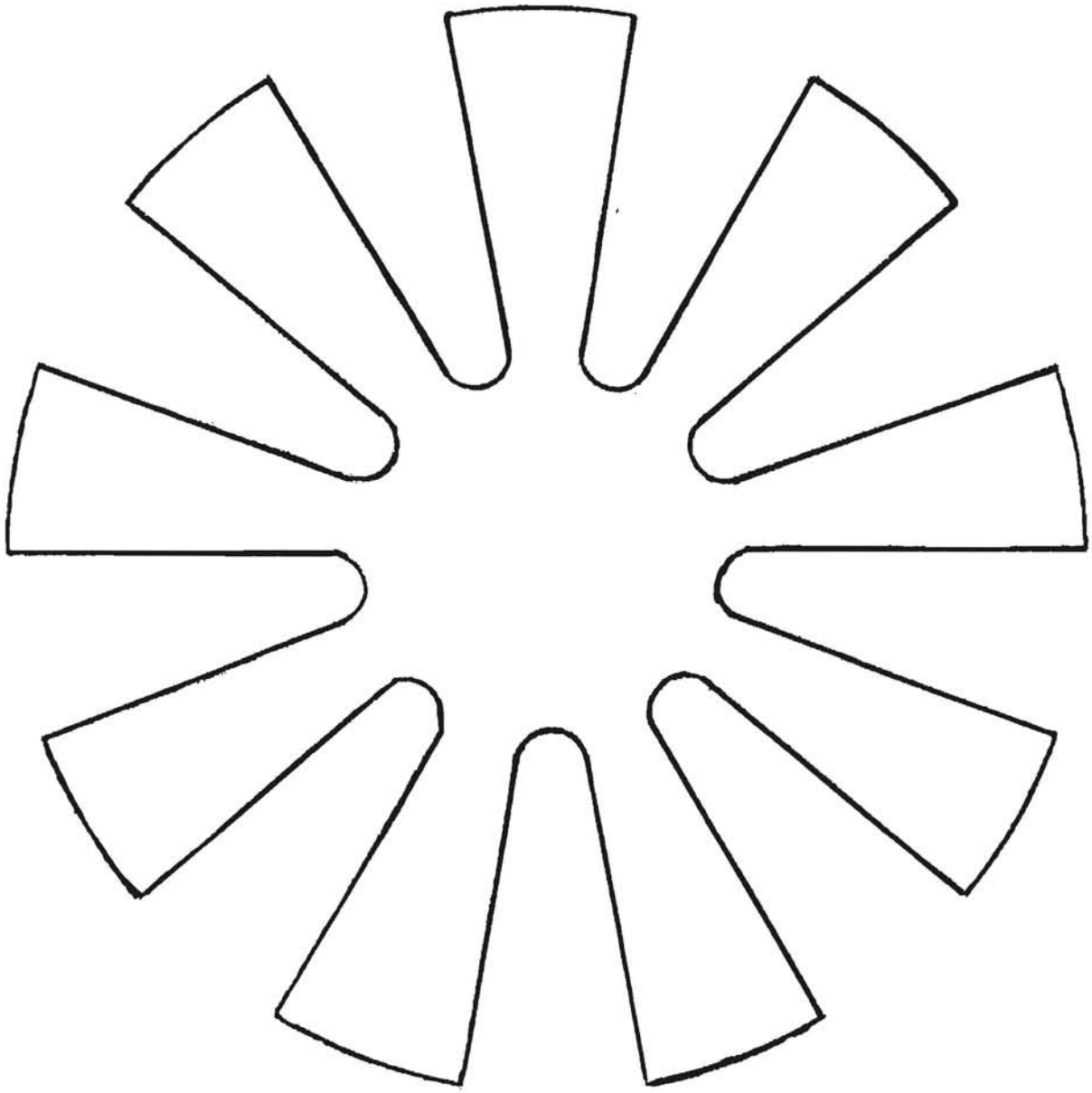
The spiderweb coil is a tuning coil that is wound on a flat wheel form that has an uneven amount of spokes. This type of coil was also used as a loop antenna in early radios.



Our receiver uses a tapped spiderweb coil that is tuned by a variable capacitor. As shown in the illustration, the spiderweb coil is mounted at the rear of a 8-inch by 6-inch by 3/4-inch wood base. A catwhisker crystal detector and a variable capacitor are mounted near the front of the wood base.

SPIDERWEB

The spiderweb coil is wound on a 6-inch diameter flat form that has nine vanes. Make a tracing of the spiderweb coil form template on the next page. Then use carbon paper to transfer the tracing to a section of 1/8-inch fibreboard or sheet plastic. Cut the coil form out with a jeweler's saw, and drill a center hole to fit a small wood screw.



Spiderweb Coil Form Template

Wind the coil with No. 22 enameled copper wire, starting from the center and winding outward. Leave about 5-inches of the starting end of the wire free. It will be used as a connecting lead. Feed the wire behind a vane and over the front of the next vane as you wind. Wind the wire closely and evenly on the form. It is not necessary to count the turns. Wind the wire to about 1/4-inch from the end of a form vane. Then drill a small hole near the vane end, and bend a solder lug through it. Solder the wire end to the lug.

On one of the vanes, count down five wires and then carefully lift it outward. Remove about 1/4-inch of the enamel with a section of sandpaper and then bend the end of a solder lug aound the bared wire. Solder the lug to the wire, then cut off remaining portion of the lug (away from the bentover end). Repeat this every five turns until you have a total of six lug taps on the coil. Stagger the lugs away from each other to keep them from shorting.

BASEBOARD

As shown in the illustration, the Spiderweb Coil Crystal Set parts are mounted on a wood baseboard. Cut the 8-inch by 6-inch baseboard from a section of 1/2-inch thick wood. White pine was used in our model. Sand the surface of the wood after cutting.

Make the spiderweb coil mounting post by cutting out a 5-inch by 3/4-inch by 3/8-inch wood section. Then cut a notch in the center of the rear of the baseboard to fit the wood section. Install the wood section in the notch with small wood screws.

Mount the parts on the baseboard as shown in the illustration with small wood screws. If the mounting holes of the tuning capacitor C1 are on the bottom, drill matching holes in the baseboard and use flathead machine screws to hold the capacitor.

Install solder lugs with the four terminals, and mount the spiderweb coil to the mounting post with a wood screw through a center hole in the coil form. The coil should be positioned with the lugs facing forward (on the left side of the coil form).

WIRING

Connect the crystal set components with hookup wire as shown in the illustration. Use 6-inch lengths of insulated stranded wire to the two small alligator clips.

Experiment 1

Connect antenna, ground, and 2000-ohm headphones to the crystal set. Then connect the D1 alligater clip lead to the antenna terminal. Set the tuning capacitor C1 to minimum capacity (rotor blades all the way out). Connect the antenna clip lead to the tap on L1 near the center of the form.

Adjust the crystal detector D1 catwhisker to a sensitive spot and tune C1 for a radio station. It may be easier to tune for stations if you temporarily connect a crystal diode (1N34A or equiv.) in place of the catwhisker crystal detector.

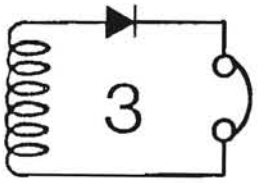
After a station is heard, you can disconnect the diode and probe for a sensistive spot on the crystal with the catwhisker. Change the antenna clip lead to other L1 taps and continue tuning C1 for radio stations.

Experiment 2

Disconnect the D1 clip lead and reconnect it to a tap on L1 close to the center of the coil. This will load L1 more lightly than if the detector was connected to the top of the coil. Repeat the tuning procedure of Experiment 1 and note the difference in selectivity and volume of the received radio stations. Then tune in a radio station and move the D1 clip lead to other taps on L1. There will be a particular tap point for optimum selectivity and volume for each received station.

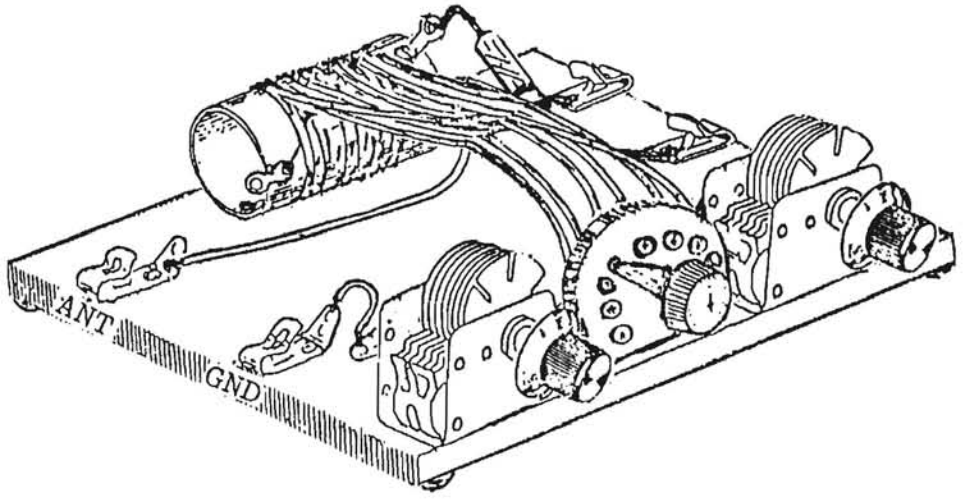
STATION LOG

DATE	TIME	DIAL	STATION	NOTES



ADVANCED RECEIVER PROJECTS

DUAL TUNED CRYSTAL SET

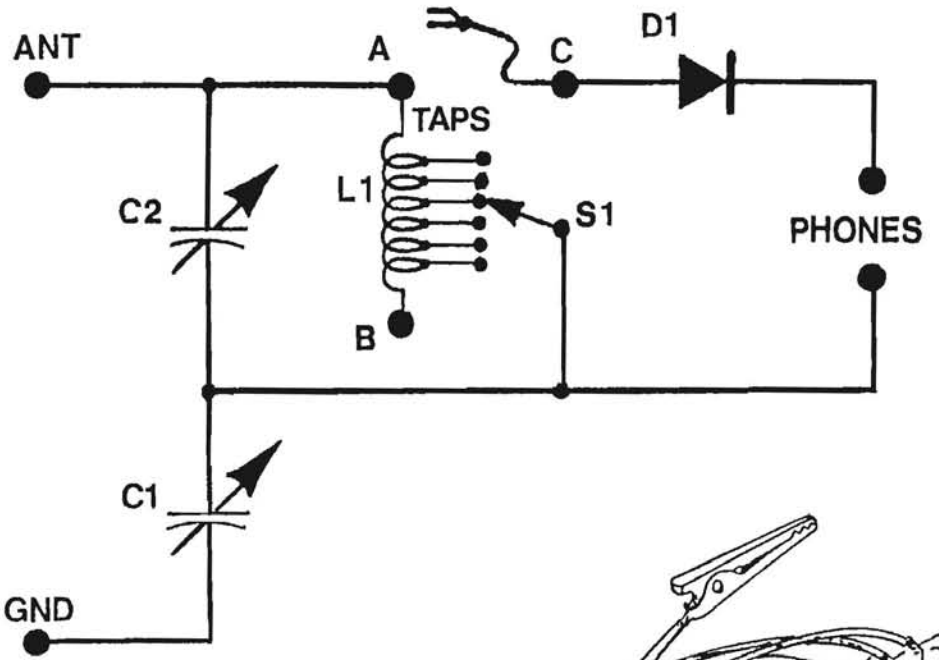


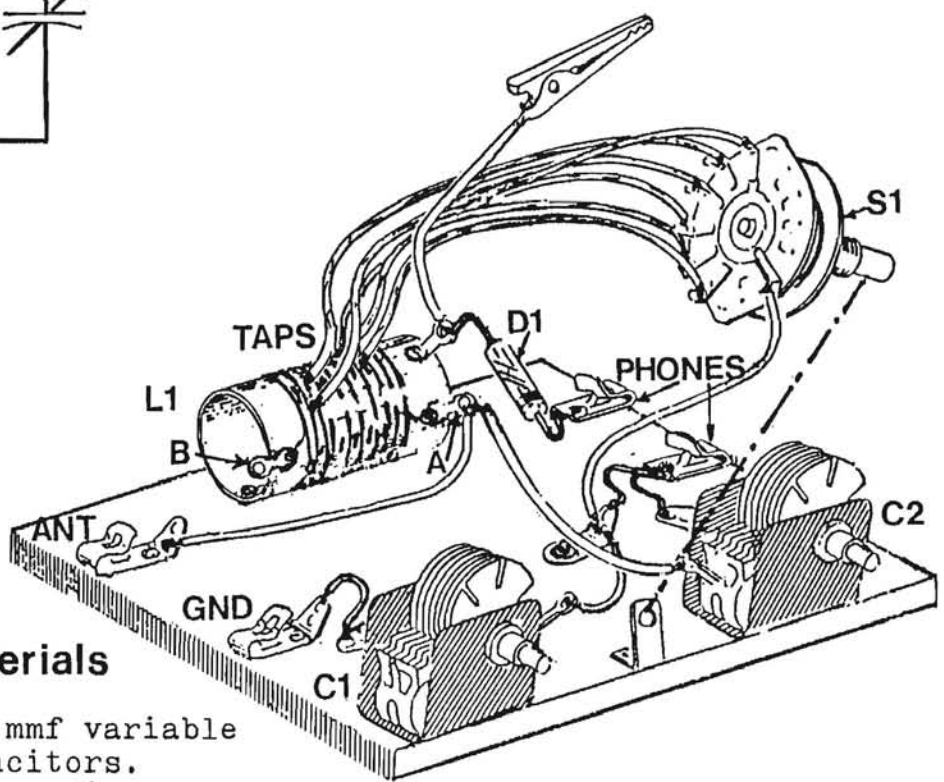
This crystal set circuit uses two variable capacitors C1 and C2. As you can see in the schematic drawing, C1 tunes the ground circuit, and C2 tunes a tapped coil L1. The taps of L1 are connected via the rotary switch S1.

TUNING COIL

Look at the drawing of the coil form construction details on page 34. The tuning coil L1 is wound on a section of 1/8-inch wall plastic tube, 2-inches in diameter, and approximately 4-inches long.

Drill holes in the form and install three solder lugs with machine screws and nuts. Position the solder lug ends towards the coil form ends as shown.





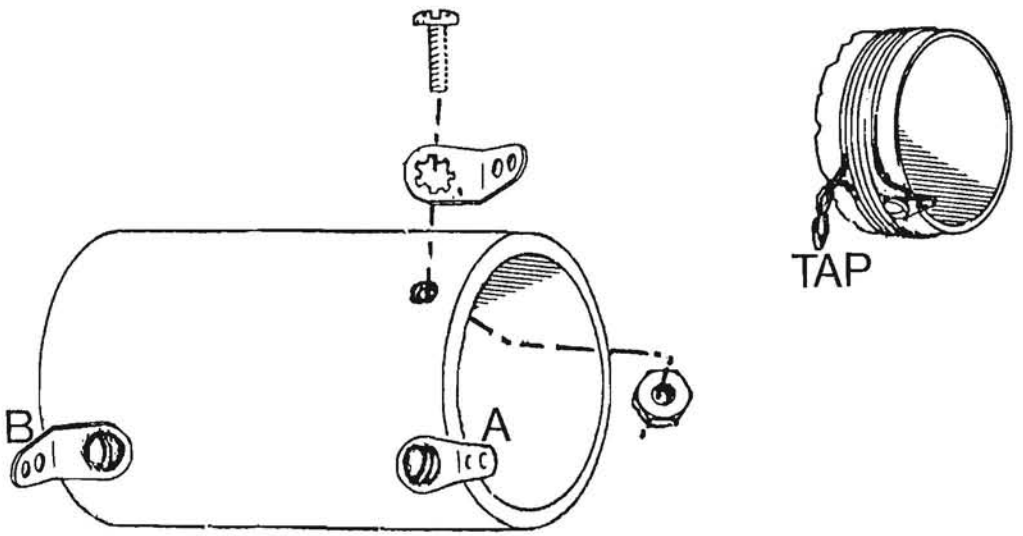
List of Materials

C1,C2 - 365 mmf variable capacitors.
D1 - Germanium diode (1N34A or equiv.).
L1 - Tuning coil (see text).
S1 - 1-pole, 10-position rotary switch (see text).
4 - Fahnestock clips.
1 - Small alligator clip.
BASEBOARD - 5 x 8 x 3/8-inch wood section (see text).
PHONES - 2000-ohm headphones.
MISC. - Coil form (see text), No. 24 enameled copper wire, solder lugs, hook-up wire, wood screws and nuts, knobs, bracket for S1.

DUAL TUNED CRYSTAL SET

The coil L1 is wound with No. 24 enameled copper wire, and is tapped every five turns for a total winding of 75 turns.

Drill a small hole in the coil form approximately 1 1/4-inch from the "A" lug. Remove the enamel from the end of the No. 24 wire and then feed the wire end through the hole. Solder the wire to lug "A".



Wind five closely spaced tight turns around the coil form. Then bend back and twist together 1/2-inch of the wire to make a tap. Position the tap outward and then continue winding the coil, keeping the turns as tight and close together as possible. Make a tap every five turns and then drill a small hole in the coil form after the 75th turn. Feed the wire through the hole, remove enamel, and solder the end to lug "B".

BASEBOARD

The Dual-Tuned Crystal Set is built on a wood baseboard as shown in the illustration. Cut the 5-inch by 8-inch baseboard from a section of 1/4-inch thick plywood. Sand the edges and surface of the baseboard to remove any splinters. Drill holes in the baseboard and install the parts with small machine screws and nuts in the locations shown in the illustration. Drill mounting holes at each end of the tuning coil L1 and position it on the rear of the baseboard with the coil taps toward the front. The illustration shows a commercial rotary switch, but a wooden rotary switch can be used for S1 instead. (See page 53 for construction details.)

Install the four fahnestock clips (used as antenna, ground and headphone terminals) with solder lugs. Mount an additional solder lug on the baseboard midway between S1 and L1.

WIRING

Connect the crystal set parts with hookup wire as shown in the pictorial illustration and the schematic diagram. Use the center solder lug as a junction point for leads from C1, C2 and S1 as shown in the illustration. Connect the leads from S1 to the taps on L1 in numerical sequence; every 5 turns for 7 taps. Then connect the last three leads every 10 taps. Insert the lead ends into the taps and hold them with a drop of solder. This will make it easier to move leads to other L1 taps. Use a 5-inch length of stranded hookup wire for the alligator clip lead connected to D1 at solder lug C.

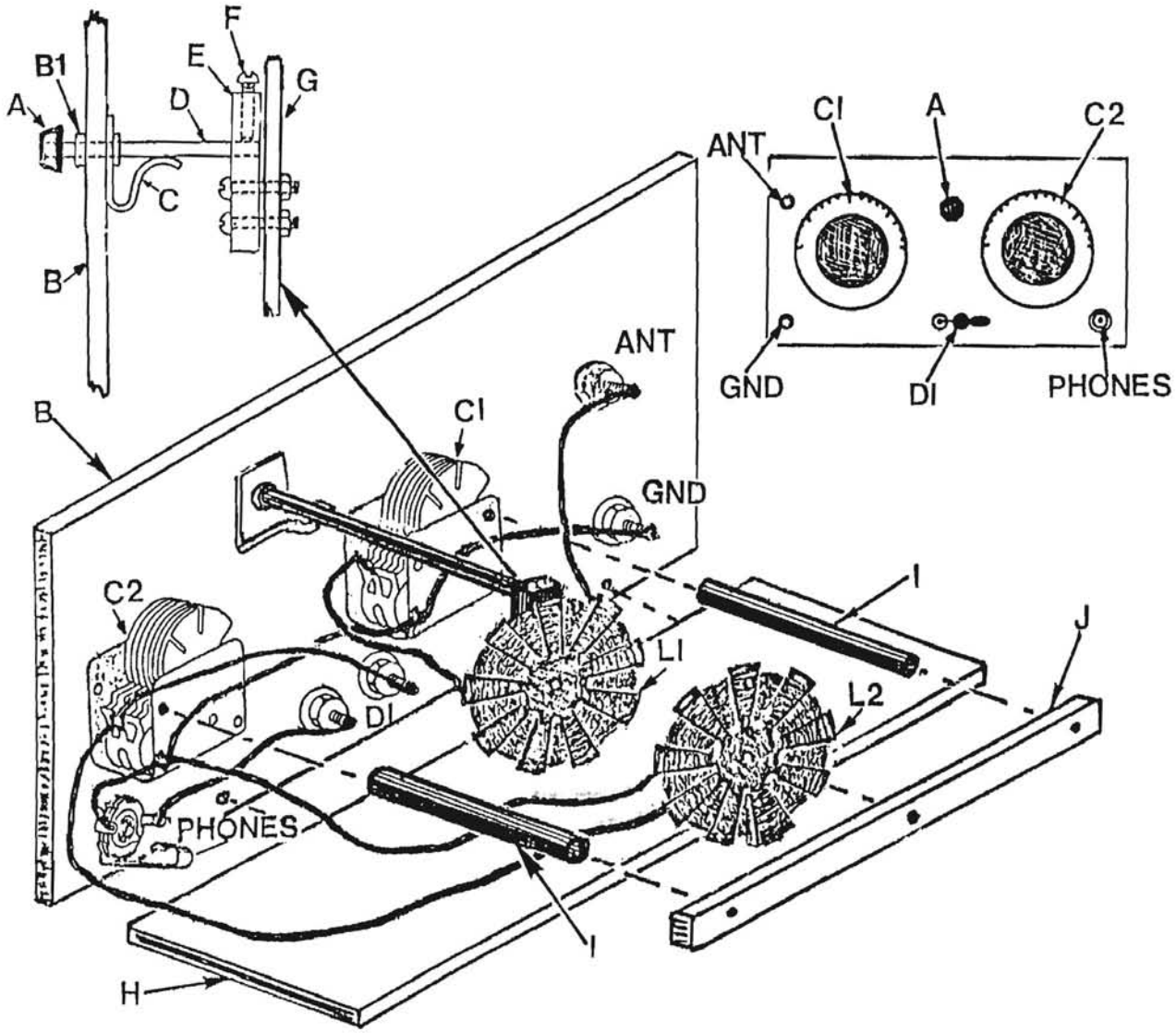
Experiment 1

Connect 2000-ohm headphones, antenna and ground leads, to the crystal set terminals. Then connect the D1 clip lead to the L1 "A" solder lug.

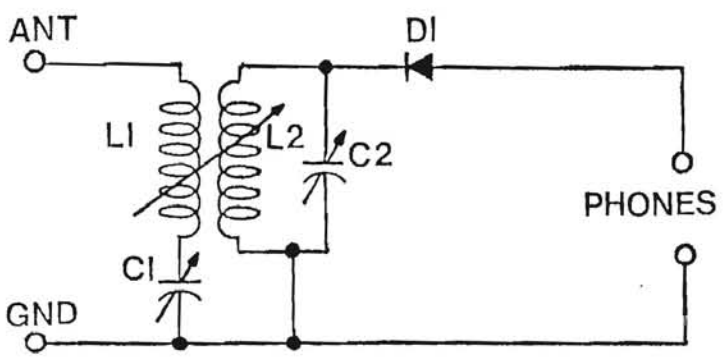
Set S1 to each tap position in turn and tune C1 and C2 for received radio stations. If all of the local stations can not be heard, experiment with different connections to the L1 taps. More turns may have to be added to L1 (with additional five-turn taps) to receive stations at the lower frequency end of the broadcast band.

Experiment 2

Remove the D1 clip lead from "A" and then connect it to the "B" solder lug on L1. Repeat the tuning method of Experiment 1 and note if the crystal set is more selective. This is possible, because L1 is now connected as an rf transformer. You can also connect the D1 clip lead to each of the S1 taps and note if there has been any improvement in reception.



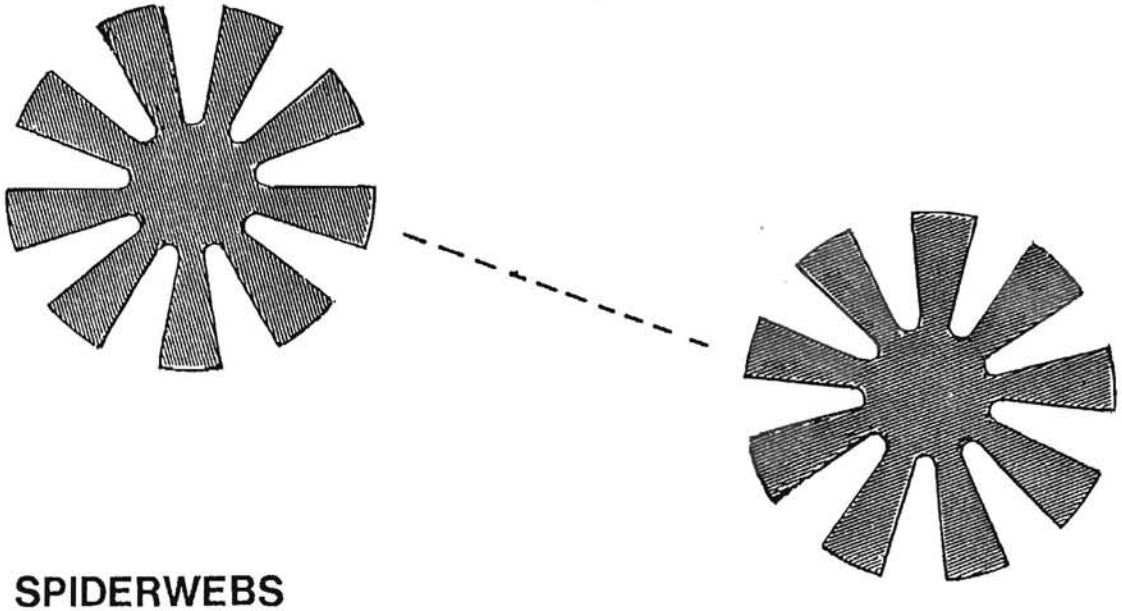
SELECTIVE CRYSTAL SET



SELECTIVE CRYSTAL SET

This receiver uses two spiderweb coils in a two-circuit tuner for good selectivity and sensitivity. As you can see in the schematic diagram, the antenna-ground rf input circuit is tuned by L1-C1 for maximum signal gain. L1 acts as the primary winding of an rf transformer, with L2-C2 as the tuned secondary. Coupling between L1 and L2 is variable for best selectivity. L2 is connected to a catwhisker crystal detector and the audio signals are fed through the phone jack to 2000-ohm headphones.

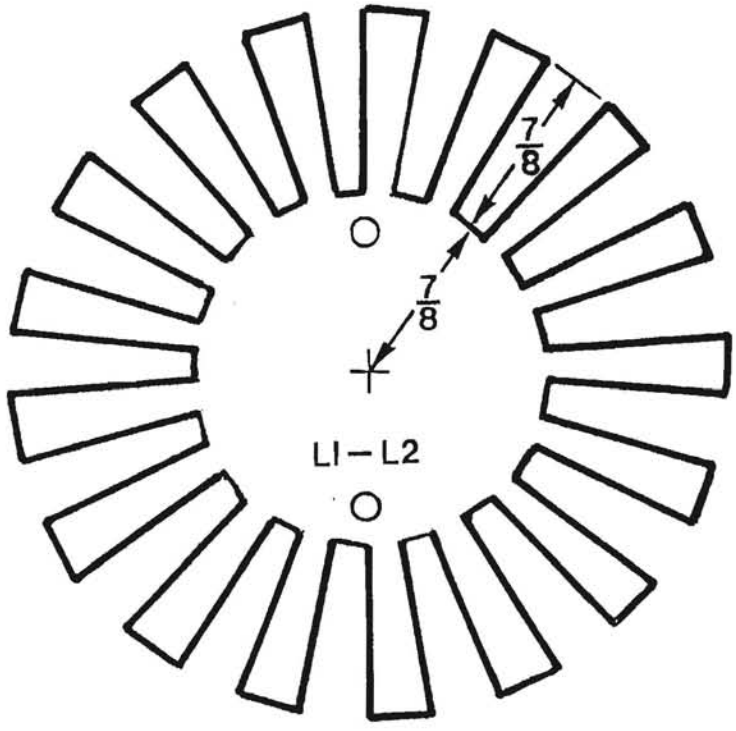
As shown in the illustration, the crystal set parts are mounted on the rear of a black plastic panel. This type of layout is similar to the type of construction of the old-time crystal sets.



SPIDERWEBS

Both L1 and L2 spiderweb coils are made exactly the same way. L1 and L2 are wound on 17-vane coil forms 3 5/8-inches in diameter. The coil forms in our receiver are made from 1/16-inch sheet plastic of the same type used for printed circuit boards (but without the copper foil). Any kind of rigid sheet plastic should be suitable for the coil forms.

Make a tracing of the spiderweb coil template. Then use carbon paper to transfer the tracing on a section of 1/16-inch thick sheet plastic. Cut out the vanes with a jeweler's saw. Drill two holes as shown, and install the two solder lugs.



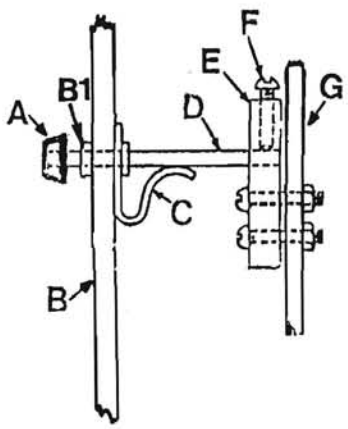
COIL FORM TEMPLATE FOR L1—L2

Wind as much No.22 enameled wire as possible on each coil form. Solder the wire to a solder lug before winding a coil. Then wind the wire over a vane and under the next vane. Wind the wire closely and evenly on the form. It is not necessary to count the turns. Solder the wire end to the remaining solder lug

CONSTRUCTION

The panel used in our receiver is a 7 3/4-inch by 4 1/2-inch section of 1/4-inch thick black acrylic plastic. The size is not critical, but it should be large enough to mount all of the parts as shown in the illustration.

Layout and drill the mounting holes of the components on the panel. Make sure that C1 and C2 are positioned so that the edges of the tuning knob dials are more than 1/2-inch apart. Install the catwhisker crystal detector assembly, antenna and ground terminals, and the phone jack in the locations shown.



As you can see in the illustration, L1 is installed on the end of a variable coupling metal rod. This 1/4-inch diameter rod is 3-inches long with a flat side. Drill and tap a small hole close to the end of the rod. The rod used in our model came from a discarded volume control. Cut a 2-inch long by 1/2-inch wide by 3/8-inch thick wood section and drill holes to fit the rod, two mounting screws, and a hole for the rod set screw. Locate and mount the wood section in the center of a spiderweb coil L1 so that the rod will be 3/4-inch down from the edge of the top vane.

Hold L1 behind the panel and using the free end of the rod, locate the panel hole. Position L1 midway between C1 and C2, and make sure that the top edge of the coil is below the top of the panel. Drill a hole to fit the rod bushing in the panel hole location. The panel bushing in our model was removed from a discarded volume control.

Drill a hole to fit the bushing in one end of a 2-inch by 1/2-inch section of sheet aluminum. Bend the other end of the section to fit against the flat side of the rod. Install the metal section and the bushing in the panel hole.

Insert the rod into the bushing and make sure that the metal section end fits close to the flat side to prevent rotation of the rod. Install a knob on the rod end and make sure that the rod can be pushed in and out easily. If necessary, put a small amount of grease on the flat side of the rod.

Install the remaining spiderweb coil L2 on a wood section that is mounted on the rear of the panel with two spacers. The spacers should be long enough for L2 to be located 1/4-inch behind L1 with the rod pushed all the way in. The 5 1/2-inch by 1-inch by 1/4-inch wood section in our model is supported with screws installed through 1 1/2-inch metal spacers into tapped holes in the rear of C1 and C2. The screws and bushings can be made longer and mount into panel holes in your receiver.

Install a wood section on the bottom rear of the panel to keep the receiver upright. A 5-inch by 3-inch section of 1/4-inch thick wood is used in our receiver.

SET WIRING

Connect the receiver components with hookup wire as shown in the illustration and schematic diagram. Use flexible stranded wire for connections to L1. Make sure that the L1 leads do not touch any component as the variable coupling rod is moved in and out.

Install knobs on C1, C2, and the variable coupling rod. Then place the crystal in the detector assembly.

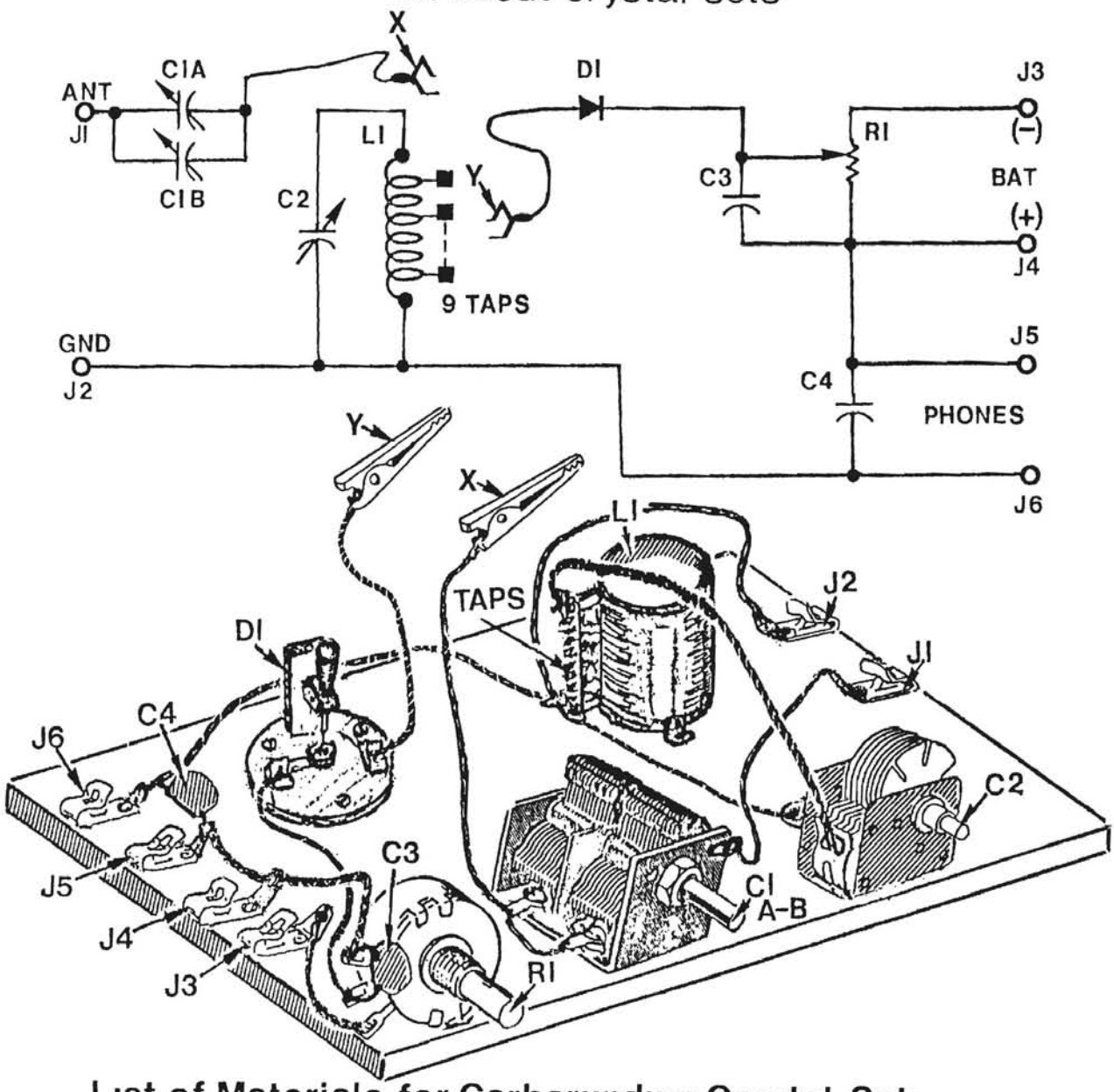
OPERATION

Connect antenna and ground leads to the receiver terminals. Plug in a set of 2000-ohm headphones into the panel phone jack. For easier initial operation of the receiver, temporarily connect a germanium diode (1N34A or equivalent) across the D1 connecting lugs. Make sure that the catwhisker is not touching the crystal at this time.

Push the variable coupling rod half way in and tune C1 and C2 for a station. Adjust the variable coupling rod and retune C1-C2 for best reception. Check operation over the entire broadcast band. It may be necessary to add or subtract turns of wire on L1-L2 for complete coverage. After checking operation with the germanium diode, remove it and try using the catwhisker crystal detector.

List of Materials

C1,C2 - 365 mmf variable capacitor.
D1 - Catwhisker crystal detector assembly.
L1,L2 - Spiderweb coils (see text).
2 - Terminals.
1 - Phone jack.
PHONES - 2000-ohm headphones.
MISC.- Plastic sheet for spiderweb coils, solder lugs, machine screws and nuts, wood screws, hookup wire.
CONSTRUCTION MATERIALS (See Text).
A - knob
B - Panel
B1- Bushing
C - Aluminum section
D - Rod
E - Wood section
F - Setscrew
G - L1 coil form
H - Wood section
I - Spacers (2)
J - Wood section



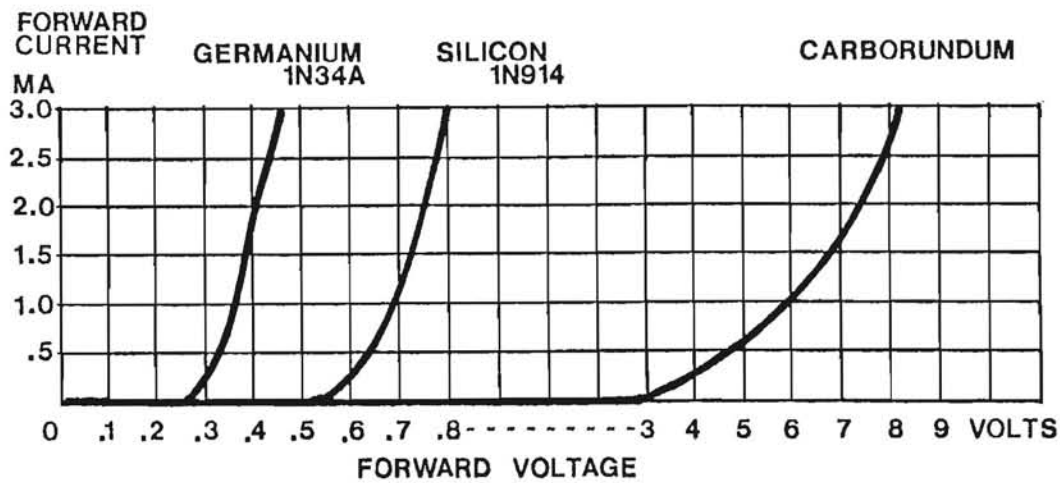
List of Materials for Carborundum Crystal Set

C1A-C1B - Dual 365 mmf variable capacitor.
C2 - 365 mmf variable capacitor.
C3 - 0.1 mf capacitor.
C4 - 1000 mmf capacitor.
D1 - Carborundum crystal and catwhisker detector assembly (see text).
J1 to J6 - Fahnestock clips.
L1 - Tuning coil (see text).
R1 - 5,000-ohm potentiometer (linear taper).
BAT - 6-volt battery
PHONES - 2000-ohm headphones.
MISC. Coil form (see text), No.24 enameled copper wire, 1/2-inch long spacers, perfboard strip (see text), push-in clips, solder lugs, knobs, hook-up wire, 1N34A diode (or equiv.), wood for base and D1 detector assembly (see text), wood screws, machine screws and nuts, alligator clips X-Y, brackets for L1 and R1.

CARBORUNDUM CRYSTAL SET

Materials other than galena, silicon or germanium have the property of detection. In the beginning of radio, carborundum (the name given to a compound of silicon carbide) was also used as a detector. The required heavy catwhisker pressure made it suitable for the early radio stations on ships, as it did not fall out of adjustment. It would also handle the strong signal energy from nearby spark transmitters without burning out. But what makes this type of crystal detector different, is the necessity for a bias battery.

A crystal detector has a high current flow with voltage applied so that it conducts in the forward direction (catwhisker to crystal), and a very low current flow in the reverse direction. The amount of current flow in the forward direction depends upon the characteristics of the crystal material and the applied forward voltage.

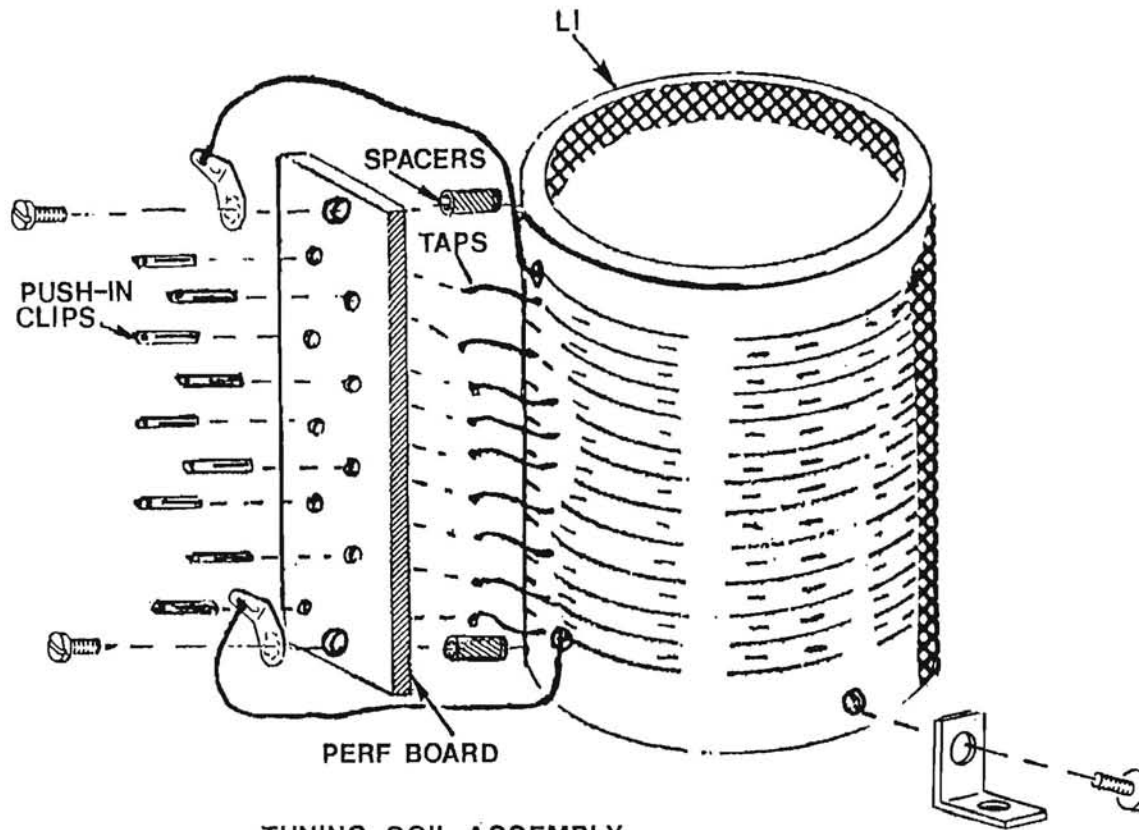


TYPICAL CRYSTAL DIODE CURRENT

The graph compares carborundum with germanium and silicon crystal detectors. Germanium minimum conduction voltage is about 0.3 V. Silicon is 0.6 V, and carborundum is 3 V. The high carborundum minimum conduction voltage is the reason a bias battery is used to move the threshold down so that weak radio signal voltages can be detected.

TUNING COIL

As shown in the illustration, the tuning coil L1 is wound on a cardboard (or plastic) mailing tube section 2-inches in diameter and 2 3/4-inches long.



TUNING COIL ASSEMBLY

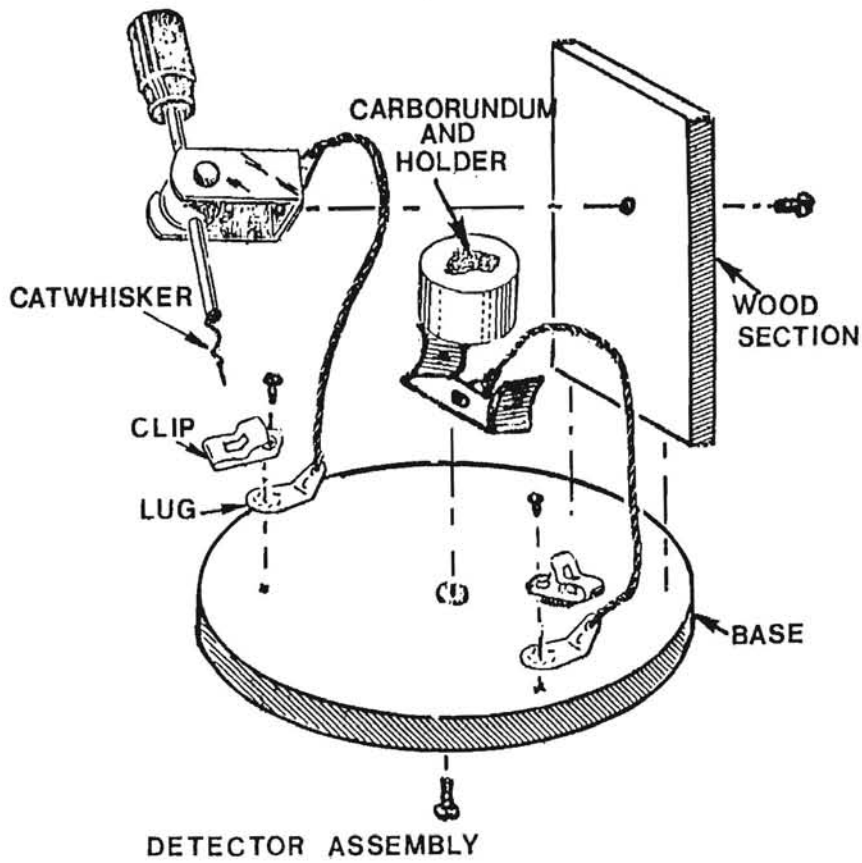
Locate and drill a hole 1/4-inch from each coil form end for machine screws to fit the two spacers. Also drill holes for two mounting brackets at one end. Then drill a small hole near each spacer hole to feed the wire on the form before you start winding.

Wind the coil with No. 24 enameled copper wire. Tap the coil every 10 turns for a total of 9 taps. An easy way to make the taps is to twist the wire together for a half inch and position the free end out. Make all the coil taps in the area between the two spacer holes. Thread the free end of the wire through the small hole at the end of the form near the spacer hole. Make sure that all of the turns are tightly wound and close together. Use a section of sandpaper to remove the enamel from the tap wire ends. Then tin the ends.

Mount 9 push-in clips in a 1/2-inch by 2 1/4-inch perf board section. Locate and drill holes on the board and install it on the coil form with two solder lugs, 1/2-inch spacers, machine screws, and nuts. Solder the coil taps to the ends of the push-in clips. Connect the coil leads to the solder lugs and mount the two 1/2-inch brackets on the coil form.

DETECTOR

The crystal detector assembly is vertical, (as shown in the illustration) instead of horizontal as in most catwhisker detectors. Begin construction by cutting the 2 1/2-inch diameter base and 2-inch by 4-inch vertical section from 1/4-inch thick wood.



Locate and install the two fahnestock clips on the base with solder lugs and small wood screws. Mount the vertical section on the edge of the base with wood screws (or cement). Locate and install the catwhisker holder with a solder lug, machine screw and nut near the top of the vertical section.

Temporarily position the crystal holder on the base so that the catwhisker will touch the center of the crystal. Locate and mount the crystal holder in this position with a solder lug and machine screw and nut. Connect the detector assembly with hookup wire as shown in the illustration. Solder a short section of coiled steel spring wire to the end of the positioning rod to serve as a catwhisker. Cut the end of the wire diagonally to give it a sharp point.

BASEBOARD

The receiver parts are installed on a 8-inch by 7-inch by 3/4-inch wood base. Wherever poossible, small wood screws are used to hold the components on the base.

Begin construction by laying out and installing the parts in the positions shown in the illustration. It will be necessary to make a metal mounting bracket to fit R1. C1A-B and C2 are mounted with machine screws in the tuning capacitor frames through countersunk holes in the base. Install solder lugs in the rear frame of both tuning capacitors. Also mount solder lugs with all of the terminals. Install knobs on the tuning capacitors and R1.

WIRING

Use hookup wire to connect the crystal set parts as shown in the schematic and illustration. Connect 5-inch leads of flexible stranded insulated wire to each alligator clip. Solder one lead to the stator lugs of C1A-B and connect the other lead to D1. Connect C3 and C4 as shown and keep their leads as short and direct as possible.

Experiment 1

Connect your antenna to J1, ground to J2, and 2000-ohm headphones to J5 and J6. Temporarily connect a germanium diode (1N34A or equiv.) in place of the carborundum crystal detector D1. The diode will make it easier to check out the receiver.

Do not attach the battery to J3-J4 at this time, or the diode may be damaged.

Connect X C1A-B clip lead to the top tap of L1, and Y D1 clip lead to the second tap from the top. Tune in a radio station with C2 and then adjust C1A-B for best volume. Change the position of the clip leads and retune C1A-B and C2 for best reception. Check operation of the receiver over the entire broadcast band.

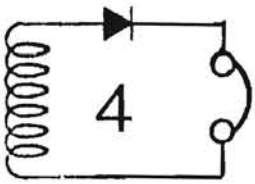
Experiment 2

Tune the receiver to a radio station and then disconnect the germanium diode. Do not disturb the receiver tuning. Install a carborundum crystal into the D1 assembly. Set R1 to mid range, and connect a 6-volt battery to the receiver; negative lead to J3, positive lead to J4.

Adjust the D1 catwhisker for a sensitive spot on the carborundum crystal. Then set R1 for best volume of received signal. Retune C1A-B and C2 and change the clip leads on the L1 taps as necessary for reception of radio stations.

Note

Carborundum that will work as a detector may be hard to find. Grinding wheels are not suitable as a source, because the carborundum is usually in fine particles mixed with other material. Best bet is to contact collectors of antique radio parts. But you can use this receiver to experiment with minerals for semiconductor operation. See the Workshop Section for information on mounting mineral sections for detector experiments.

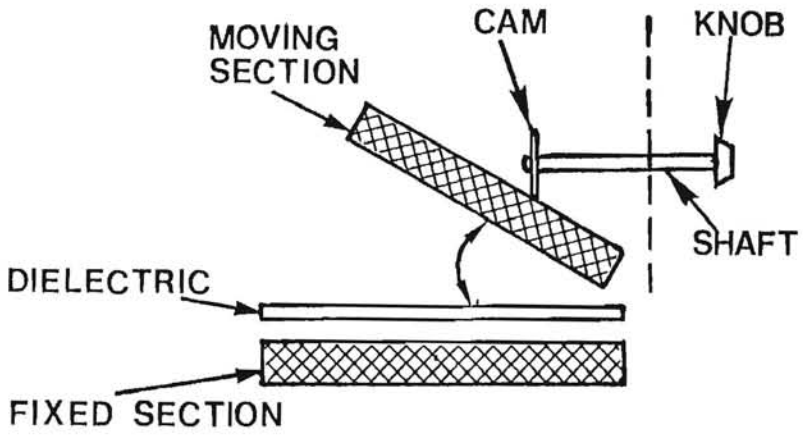


COMPONENT PROJECTS

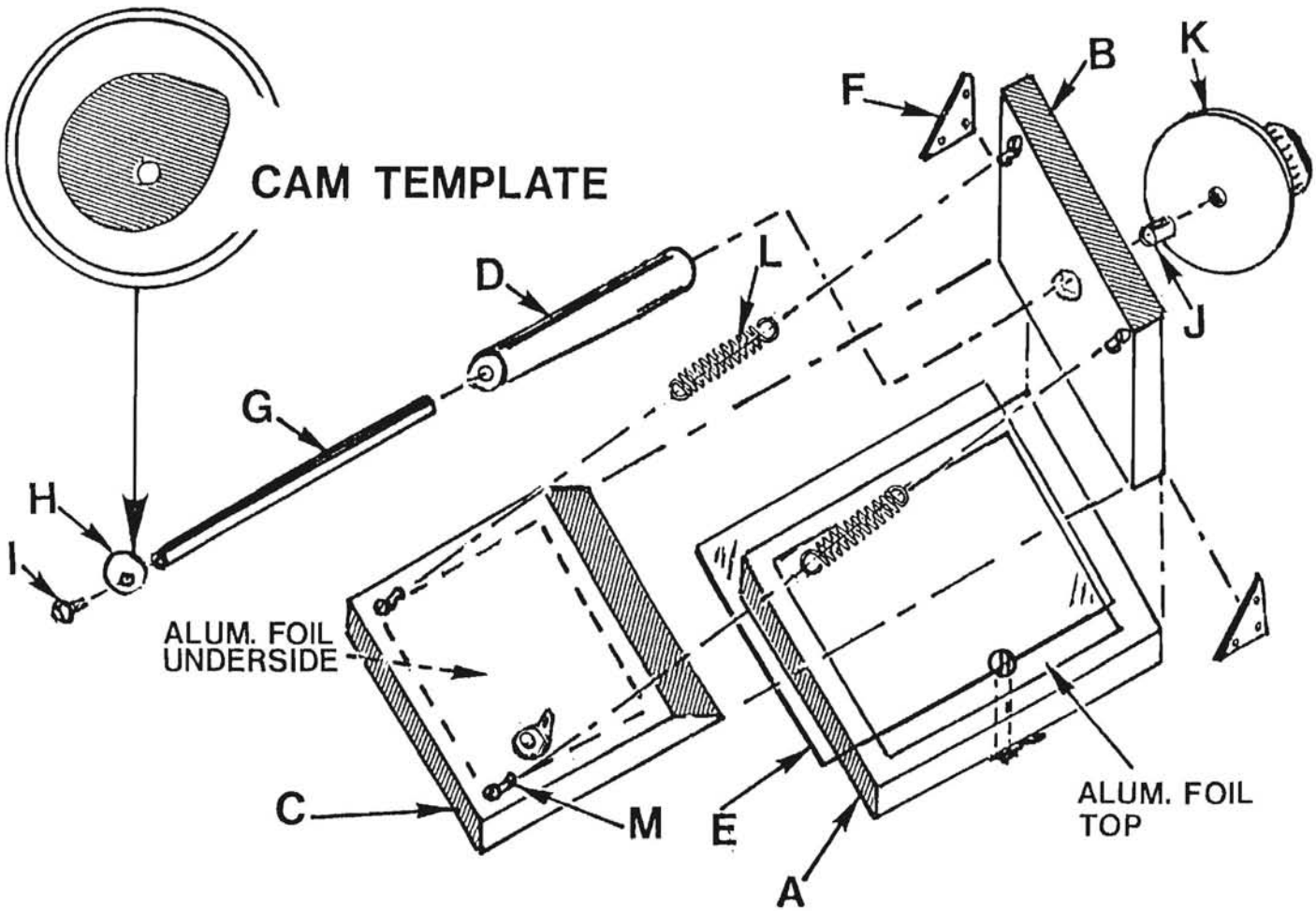
BOOK TUNING CAPACITOR

During the roaring 20's,the Crosly Radio Corporation used a type of tuning capacitor that had two plates hinged together on one edge. The dielectric was a sheet of mica between the two plates. This type of tuning capacitor was called a Book Condenser.

The tuning capacitor shown in the illustration is similar to the Crosly Book Condenser. It uses aluminum foil plates cemented to fixed and movable wood sections, and a plastic sheet dielectric (cut from thin transparent plastic notebook protector material).



Adjusting the tuning knob causes a metal cam to rotate and control the movement of a spring loaded wood section. This varies the capacity between the two foil plates. As the plates are brought closer together, the capacity increases. As they are separated, the capacity decreases.



BOOK TUNING CAPACITOR

List of Materials

A - 3 x 3 1/2 X 3/8-inch wood section
B - 3 x 1 3/4 x 3/8-inch wood section
C - 3 x 3 1/8 x 3/8-inch wood section (one end beveled 45 degrees - see text).
D - 3/4 - inch diameter wood dowel, 2 1/4 inches long.
E - 3 1/2 x 3 -inch sheet plastic (See text).
F - 2 triangular 1-inch metal sections
G - 1/4-inch metal rod, 3 -inches long.
H - Metal cam (see text).
I - Machine screw (see text)
J - Bushing 3/8-inch long (to fit rod G).
K - Tuning knob (to fit rod G).
L - 2 springs 1/4-inch diameter by 1 1/4-inch long.
M - 4 solder lugs.
MISC. Small wood screws, nails, aluminum foil, machine screws, nuts, solder lugs and washers.

CONSTRUCTION

Start construction by cutting the wood sections from 3/8-thick wood to the sizes shown in the drawing. Section C has a 45 degree bevel at one end. This acts as a pivot point for the movable section. Drill and countersink a hole to fit a flathead machine screw in the places shown on sections A and C. Do not install the screws at this time.

Cement aluminum foil on the top of section A and on the bottom of section C. Then cut and remove a 1/2-inch strip of foil from the front edge of section A. Cut a 1/8-inch strip from the remaining edges. Also cut away a 1/8-inch strip from all the edges of the foil on section C.

Carefully slit the foil on top of the machine screw hole on section A and press the foil gently inside the hole. Install the flathead machine screw into the hole, making sure that it is in contact with the foil. Fasten the screw with a washer and nut. Make sure that the screw does not turn and rip the foil. Install a solder lug and nut on the free end of the screw. Repeat this operation with section C.

Cut the sheet plastic E to size and place it on top of section A. Then place section B on top of the plastic sheet and install it on the end of section A with wood screws. Install the two metal sections F with small nails.

Saw the metal rod G to size and tap one end to fit a machine screw. Cut the dowel D to size and center drill it to fit the rod G. Cut a flat on one side of the dowel.

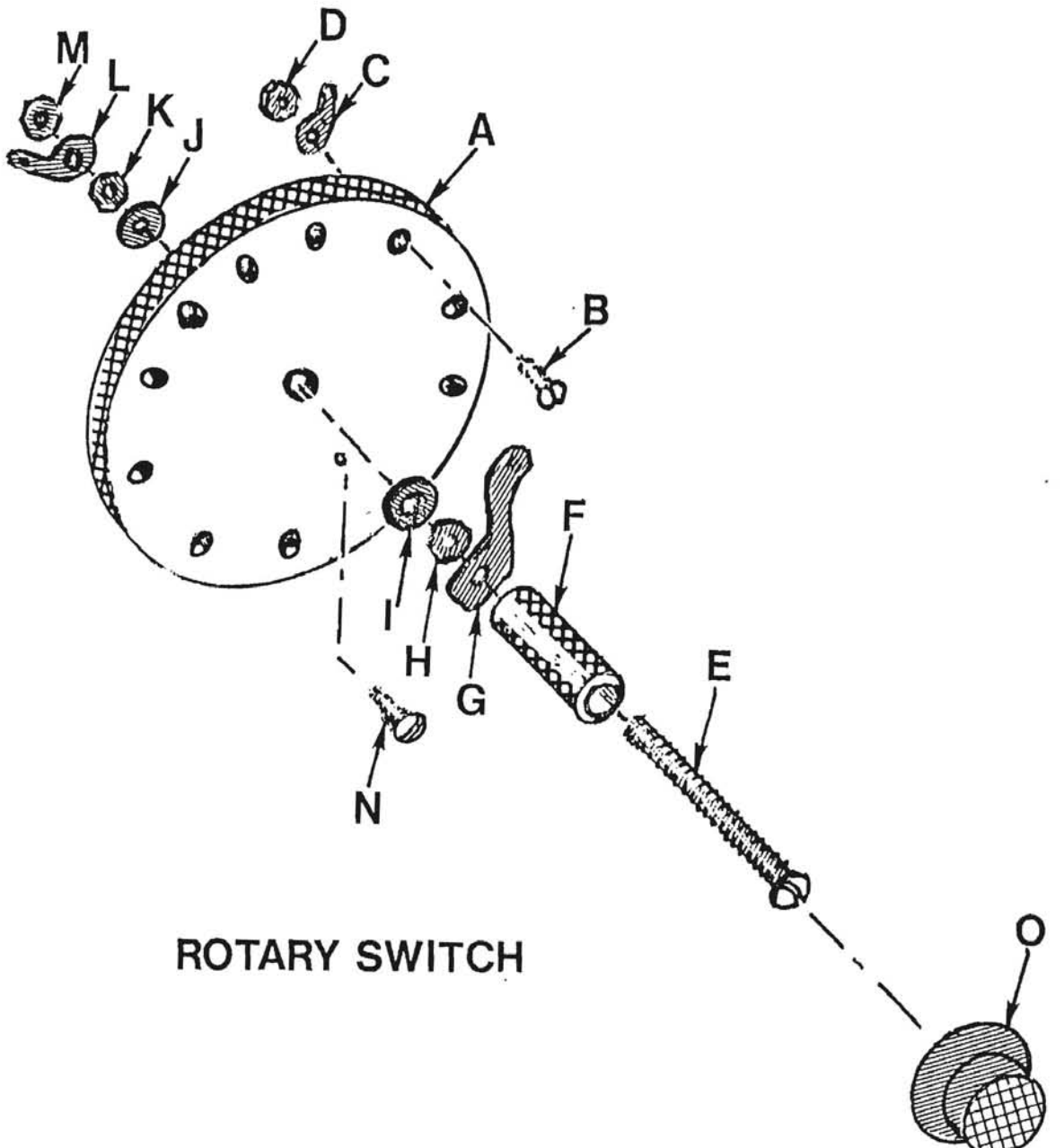
Make a tracing of the cam H template and cement it on a section of heavy gauge sheet aluminumm. Cut the cam to size and drill a hole to fit the rod machine screw in the location shown. Install the cam with a lock washer on rod G. Insert the free end of the rod into dowel D. Place section C on top of the plastic sheet E on section A. Then temporarily position dowel D with rod G in the center of section B. Rotate the cam so that its long end touches the top of section C. Mark the placement of the dowel on section B. Drill a hole in section B to fit dowel D. Cement the dowel in place.

Locate and install solder lugs with wood screws on sections A and C as shown in the drawing. Bend over the ends of the solder lugs and install two springs L in the lugs.

Mount tuning knob K and bushing J on the free end of rod G. Adjust the knob and check that section C moves in and out with the cam rotation. changing the capacity between the foil plates.

You may have to experiment with the cam diameter for best operation of section C. File the cam edges in small increments only and recheck the operation often. Use very light springs to keep section C from pressing against the cam and moving it.

Reverse the cam on rod G to alter the direction of capacity (clockwise or counterclockwise). The capacitor can be installed vertically on a receiver project with a small metal bracket mounted on the bottom side of section A.



ROTARY SWITCH

List of Materials

A - 2 1/4 diameter x 3/8-inch wood section.
B - Flat head machine screws.
C,L - Solder lugs.
D - Nuts to fit B.
E - Machine screw 1 1/4-inches long.
F - 1/4 diameter x 1/2-inch long bushing (to fit E).
G - Switch arm 1 x 1/4-inch (sheet brass, see text).
H,K,M - Nuts to fit E.
I,J - Washers to fit B.
N - Small woodscrew.

ROTARY SWITCH

In the early days, radio experimenters could not buy parts that are readily available today. They had to build their radio components from wood sections and common hardware.The rotary switch ilustrated is similar to a type constructed for use with tapped coils of crystal sets.

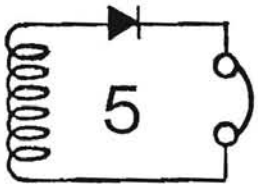
The rotary switch shown uses flat head screws for contact points that are mounted on a wood section. The switch arm is cut from sheet brass and is installed on a machine screw.

CONSTRUCTION

Refer to the drawing of the rotary switch. Begin construction by cutting the 2 1/4-inch diameter section A from 3/8-inch thick wood. Locate, drill,and countersink the holes for the flathead machine screws B. The rotary switch shown has 10 machine screws, but you can space the hole locations for the number of switch contact points desired. Allow a space between the starting and ending contact points for the woodscrew N. This woodscrew is used as a stop for the rotary arm G. Locate and drill a center hole for the machine screw E. Install the screws B in section A with solder lugs C and nuts D. Install the woodscrew N.

Measure and cut the rotary arm G from a section of sheet brass. The arm should be long enough to reach the contacts. Make sure that it is not too wide, or it may short out to a nearby contact. Make a dent or "dimple" on the contact end, and drill a hole for screw E on the other end. Bend the rotary arm as shown in the drawing. Install screw E into bushing F, and arm G, and tighten the assembly with nut H. Place washer I on the screw and install screw E into the center hole of A. loosely fasten the screw with washer J and nut K. Then rotate the bushing F and arm G. Adjust the tightness of nut K as required to allow rotation of G with only a slight "drag". Install solder lug L on screw E with nut M. and bend the lug towards the screw end. Install knob O on bushing F, and recheck operation of the arm G. It should touch all of the contacts B and stop at screw N.

The rotary switch is used as shown in the Dual-Tuned Crystal Set Construction Project. Make sure that extra lead length is allowed for connection to the solder lug L. Bend the lead into a "pig-tail," so that it will not break as the solder lug L is rotated.



WORKSHOP

The construction projects in this book can be built with common hand tools found in an average home workshop.

Useful Hand Tools

Safety glasses or face shield
Steel rule
Square
Hack saw
Wood saw
Screwdrivers
Pliers
Wirecutters
Tin snips
Center punch
Files
Hammer
Soldering Iron

Wear safety glasses or face shield while working. Observe common sense safety precautions.

SOLDERING

Make sure that the part to be soldered is clean. Use only rosin core solder. Prepare the soldering iron tip as follows:

File each surface of the tip to remove oxidation. Plug in the iron and let it heat until you can melt solder on the tip (before the copper tip turns blue or bronze color). Melt enough solder until all sides of the tip are coated. Then wipe off excess solder with a damp sponge or cloth.

Make sure that the bare wire lead to be soldered is free of oxide. It is best to tin it first by melting a bit of rosin core solder on the lead end. Crimp the lead in the part hole to keep it from moving during soldering. Hold the iron tip under the part until it is hot enough to melt solder. Apply only enough rosin core solder to cover the surface of the part. Remove the iron. Do not move the part or wire while it is cooling.

PROJECT CONSTRUCTION TIPS

Tape a sheet of graph paper over a project wood section. Mark locations of the parts on the paper. Center punch mounting hole centers through the paper. Use a compass to draw circles the same size as the holes. Then drill holes through the paper.

Lettering on the projects can be done with a labelmaker that prints on strips of adhesive coated plastic. You can also use decals or dry transfer lettering sheets.

Small metal brackets can be made from heavy gauge sheet aluminum. Layout and drill the mounting holes and then cut it to size with a tin snips. Form the bracket by using a heavy mallet and a vise.

Tuning Capacitors, Knobs, and Dials

Old radios are a good source of components. The radios generally use two tuning capacitor sections ganged together. The receivers previously described will operate with one of the capacitor sections (365 mmf). You can cut out cardboard discs and cement them to radio knobs and use them as dials with the tuning capacitors. Mark the dials with dry transfer lettering, or black drawing ink.

Panels and Cabinets

The layout and construction of crystal sets is not critical. They can be fitted into many different types of wood cabinets. You can use almost any type of non-metallic box, but metal boxes are not recommended as sheet metal close to the tuning coil may lower its "Q". Black sheet plastic used for the front panel with white dry transfer lettering will give your receiver project an antique look.

MOUNTING MINERALS FOR CRYSTAL DETECTORS

Minerals (galena, etc) for experimentation as crystal detectors, will work best when mounted in a metallic base. For temporary mounting (while testing), wrap aluminum foil tightly around the sides and bottom of the mineral. Then place it into a sheet metal ring or cup that will fit into the detector assembly.

After testing, the mineral can be permanently mounted in a metal slug. Drill a shallow hole of a diameter equal to the detector assembly crystal in a block of wood. Fill the hole with melted solder. While the solder is still molten, carefully place your mineral section into the center. Hold the mineral in place with tweezers or long nose pliers until the solder cools. Then pry the metal slug out of the wood. Be sure that the top of the mineral section is free of solder.

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About the Author

Charles Green has been involved in the design and construction of electronic equipment for industry and the U.S. Government for 30 years. Much of his experience has been as a technical writer preparing operation and maintenance manuals. His continuing interest is in constructing radios and he has had over a hundred articles published in hobbyist magazines over the past two decades.

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CHAPTER 1

HF ANTENNAS

Balcony Antenna

by **Harry Lythall - SM0VPO**
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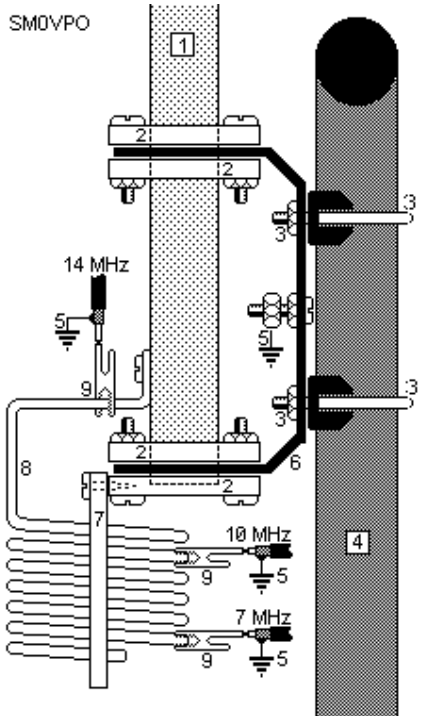
Credit Line: http://web.telia.com/~u85920178/

Many amateurs are very restricted with the space they have available for HF antennas. I have documented a short antenna for the HF bands, but here is a simple method of mounting it, and a method of further reducing the physical length. I used to use an old CB (27 MHz) half-wave antenna which had a broken matching coil. this I used as a 1/4 - wave antenna for 14 MHz, after removing the matching coil. Today I find that CB antennas have increased in price, so I have found a cheap replacement that can be fitted to the balcony of apartment dwellers.



THE ANTENNA

Above is the side view of a bracket, which can be thrown together in a couple of hours and gives surprisingly results. I use six sections for the antenna itself, each of which is 1 meter long. Each section fits inside the previous section by exactly 10 cm. the last section is adjusted so that total length of the antenna is 5.35 meters. This resonates at 14.1750 MHz. I used the following aluminum tubes:-

section 1 : 31 mm Dia. Wall thickness = 2.0 mm. (bottom section)
section 2 : 25 mm Dia. Wall thickness = 2.0 mm.
section 3 : 20 mm Dia. Wall thickness = 1.5 mm.
section 4 : 15 mm Dia. Wall thickness = 1.5 mm.
section 5 : 10 mm Dia. Wall thickness = 1.5 mm.
section 6 : 6 mm Dia. Wall thickness = 1.0 mm. (top section)
This is shown as item (1) in the drawing above.

THE BRACKET (6) & (2)

The bracket screws on to a handrail of the balcony. In my present situation I have a 7 meter wide terrace with a horizontal handrail, but there are four vertical steel pipes supporting the handrail. The bracket is screwed on to one of these vertical supports (4). The bracket is formed using 3 - 4 mm thick aluminium plate (6) with a 50 mm hole in the center of the top & bottom ends. Bend the plate in two places to prevent the plate becoming weakened. The two ends are each sandwiched in between two nylon blocks (2). Use a chopping board stolen from the kitchen, if you can get away with it. Otherwise, the chopping boards are available from:
IKEA (Sweden)
WOOLWORTHS (UK)
SAFEWAY (USA)

Drill THREE of the nylon blocks, in the center, to fit the 31mm tube (1). The fourth block (bottom) should be drilled with a 5 mm hole to allow water to run out. The bracket is bolted to the balcony handrail, using 35 mm exhaust (muffler) clamps (3).

THE COIL (7) & (8)

This is used to make the antenna resonate at lower frequencies. I wound all my coils using 4mm aluminum wire, but copper hydraulic brake pipe works as well. The coil is 10mm Dia (the same as a tin of DelMonte pineaple chunks)! The coil pitch is 1cm per turn. I used two pieces of plastic conduit (7) to support the coil.

The coil uses about 1 meter of wire/pipe for every three turns. Flatten one end and drill a hole in it for connecting it to the antenna pole (1). If you use aluminum wire, then shorter pieces can be joined together with a brass insert from a car cable connector. Copper tube can easily be soldered.

FEEDING THE ANTENNA (5) & (9)

Feed the antenna with 50 ohm coaxial cable, braid connected to the bracket (5) and the center conductor connected to an aligator clip. Select the band using the aligator clip (9):-

0 turns = 14 MHz (20 meter band) (VSWR - almost 1:1)

2 turns = 10 MHz (30 meter band) (VSWR - almost 1:1)

6 turns = 7 MHz (40 meter band) (VSWR - about 1.1:1)

51 turns = 3.8 MHz (80 meter band) (VSWR - about 1.4:1)

53 turns = 3.7 MHz (80 meter band) (VSWR - about 1.4:1)

55 turns = 3.6 MHz (80 meter band) (VSWR - about 1.4:1)

57 turns = 3.5 MHz (80 meter band) (VSWR - about 1.4:1)

Here is a photograph of one of the prototypes in my balcony. In the background you can just make out another one of these antennas, but with a bigger (63 turn) coil.

OTHER INFORMATION

The mounting is very rubust, yet the wind resistance is rather low. Both my antennas have stood up to gale force winds; they hardly wobble!! You do not have to use 5.35 meters of for item (1) if you want to work on other bands, such as 18 MHz.

If you intend to use more than 10 watts, then make sure you have a good
1 cm, or more, of insulation between the aluminium pole (1) and the bracket (6).

Have fun with this project. Regards from Harry - SM0VPO **-73!-**

Balcony Antenna Extension

by **Harry Lythall - SM0VPO**
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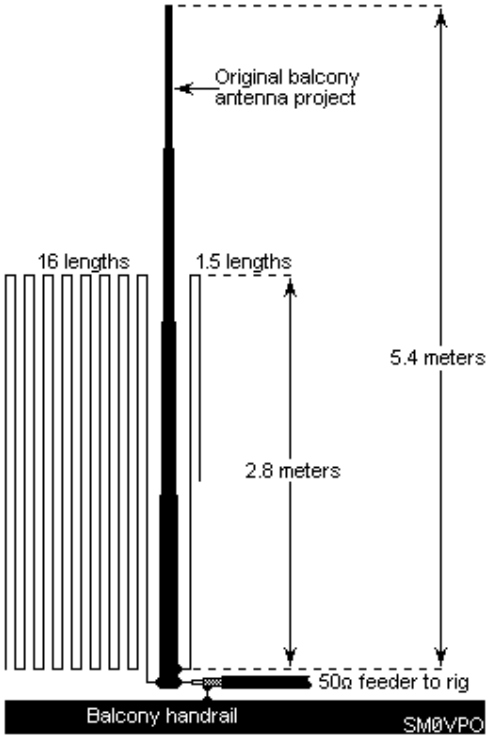
http://web.telia.com/~u85920178/antennas/balcant2.htm

You may have already seen my HF Balcony Antenna which was designed solely for 14MHz, then a coil was added to cover all the lower HF bands (10, 7 and 3.5MHz). Following an article in RadCom I have now extended this antenna to cover all bands from 3.5MHz through to 30MHz without any switching or tuning. The antenna functions using both Fractal and Meander principles. The height of one turn of the loop gives coverage of the 10-meter band, the old balcony antenna covers 20-meters, an extra element covers 17-meters and the 40-meter long meander gives coverage on the 80-meter band. Here is the measured range of the complete prototype antenna:

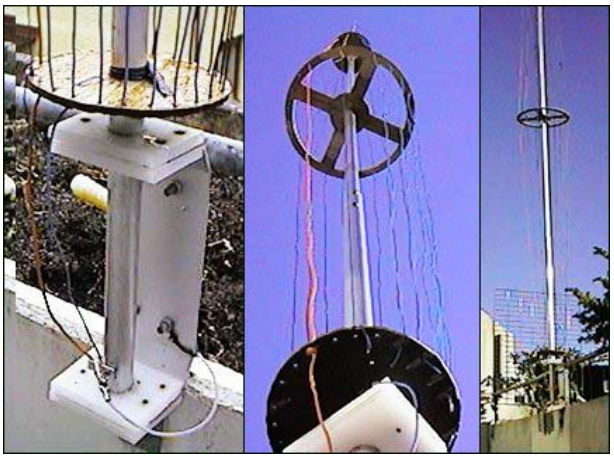
Band	Range (MHz)	Worst VSWR	Center VSWR
80 m	3.55 - 3.70	3:1	1.1:1
40 m	7.00 - 7.10	2.2:1	2.2:1
30 m	10.10 - 10.15	2.3:1	2.3:1
20 m	14.00 - 14.35	1.1:1	1:1
17 m	18.07 - 18.17	1.2:1	1.2:1
15 m	21.00 - 21.45	2.8:1	2.5:1
12 m	24.89 - 24.99	2.1:1	2.1:1
10 m	28.00 - 29.20	3:1	1.1:1

As you can see, the VSWR rises on some of these bands but the antenna is still 100% useable on all

bands without an ASCTU (ASTU or ATU). I have not tested the coverage outside amateur bands, I stopped when the VSWR became 3:1 or when the band edge was encountered. So what is the big secret? I have mentioned before in these pages that several 1/4-wave or 1/2-wave antennas can be placed in parallel and fed with a single feeder. The resonant element will have an effect; the others presenting a high impedance. I tried to add two 1/4-wave antennas to cover the original 14MHz plus 29MHz, 18MHz and 3.6MHz. When I tried it I was surprised that the antenna covered as much as 200KHz of the 3.5MHz band and other HF bands were ALL useable. Reports suggest that the effects on 14MHz have introduced a couple of dB's loss, but that is far less than one "S-point". Here is the drawing of the antenna showing the original 14MHz pole (center) and the other two 1/4-wave antennas I have added.



I have shown a graphic likeness of the routing of the additional 43-meters of wire, they are wound on three plywood disks. The top and bottom disks are 100mm diameter and the center spacer disk is 300mm diameter, each drilled with 18 holes. It would have been better to have used nylon food reparation boards (from Ikea) but I didn't really have all that much confidence this antenna experiment would work so I began with this make-shift arrangement. I must also point out that putting your hand near this antenna will cause changes to the readings, so you may need to make a few minor adjustments in your own individual case. Here are photographs of the finished and working prototype antenna.



The left insert shows the antenna mounted on the old balcony support bracket with the coil removed. The center insert shows a view from the bottom of the antenna. The orange wire is the 420cm 18MHz element. The right insert shows most of the complete antenna from a little distance. Notice how I have cut out material from the center spacer to reduce wind resistance and to help make it look a little less obtrusive for neighbors. The top spacer is identical to the bottom spacer. All three of the elements are connected in parallel at the feed point where I connected my 50-ohm feeder. The old coil is now obsolete and has been removed.

Please note that this antenna idea is also governed by "Harry's Law" of coils:
You cannot wind coils like me and I cannot wind coils like you.
Coil-winding data is a constant that varies from person-to-person.
This means that it may NEED some adjustment in your own environment, depending upon proximity of other artifacts, humidity, groundplane efficiency and even the color of the flag you have fitted to the top of the original 14MHz pole.
Begin antenna assembly by making and fitting the top, middle and bottom spacers. To trim the spacers, temporarily add a 3-meter length of wire, making a small tight loop at the top to remove the surplus. Check the VSWR at 29MHz and adjust the top-spacer position, re-coiling the surplus wire, until the antenna is resonant with VSWR better than 1.5:1. Fix the spacer positions using hose-clamps or whatever other bright ideas you may have. Now remove the 3-meter wire and sew the 40-meters of wire through the holes. Check the VSWR at 3.6MHz, or whatever part of the 80-meter band you want. Remove wire to achieve resonance. Fit the 1.5-loop, 4.2-meter length of wire for the 18MHz element. This loop only comes 1/2-way down the cage, so add some nylon line and secure it to the bottom spacer. Do not tie anything to the center spacer. The wire I used was 7-ampere multi-strand household mains-wiring cable.
Have fun with this project. Regards from Harry - SM0VPO -73!-

MULTIRANGE VERTICAL ANTENNAS

Igor Grigorov, RK3ZK
antentop@mail.ru

A combined three-band antenna

Three band antenna fundamentals: At a lack of the place for installation of a separate vertical antenna for each of three upper HF ranges it is possible to use a combined three-band antenna that works at the ranges itself. **Figure 1** shows schematic of a combined three-band antenna.

Figure 1 A combined three-band antenna

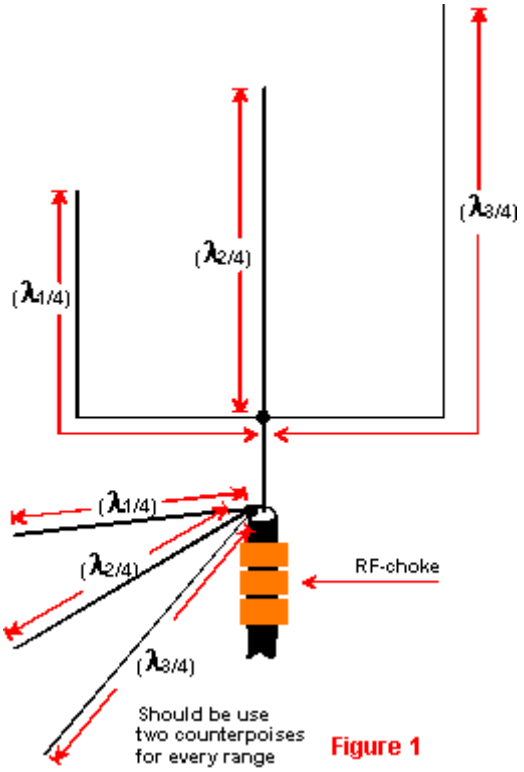


Figure 1

The antenna consists of from three quarter-wave verticals that are resonated for each of working ranges. The verticals are connected in the bottom together. Two quarter-wave counterpoises should be use for each operation range of the antenna A coaxial cable with 50-Ohm characteristic impedance will do well for the antenna. A coaxial cable with 75-Ohm characteristic impedance also would be work with the antenna, but a SWR in the coax will be higher compare to 50-Ohm coaxial cable. **Table 1** shows the combination of ranges where a mutual influence of vibrators against each other is minimum.

Design of the Antenna: Three various designs of the three- range antenna are shown below. **Figure 2** shows a simple design suitable for 6 - to 15-M. The three vibrators are placed on a small distance from each other. The distances between the vibrators are fixed with the help of small plastic insulators. The design has very strong mutual influence for every vibrator against each other.

Table 1

6m	10m	15m
10m	15m	20m
12m	17m	30m
15m	20m	40m
15m	17m	20m
20m	30m	40m
30m	40m	80m
40m	80m	160m

Figure 2 Simple design of a three ranges antenna

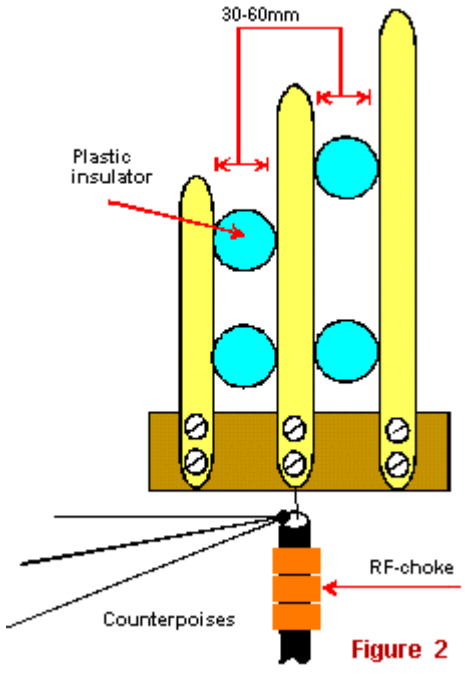


Figure 2

Figure 3 shows a simple design suitable for 6 - to 17-M. Antenna has the triangular shape. Special 'sitting' should be used for the antenna design. Vibrators are screwed in the bottom with the help of strong screws. The design has a small mutual influence for every vibrator against each other.

Figure 3 A triangular shape antenna design

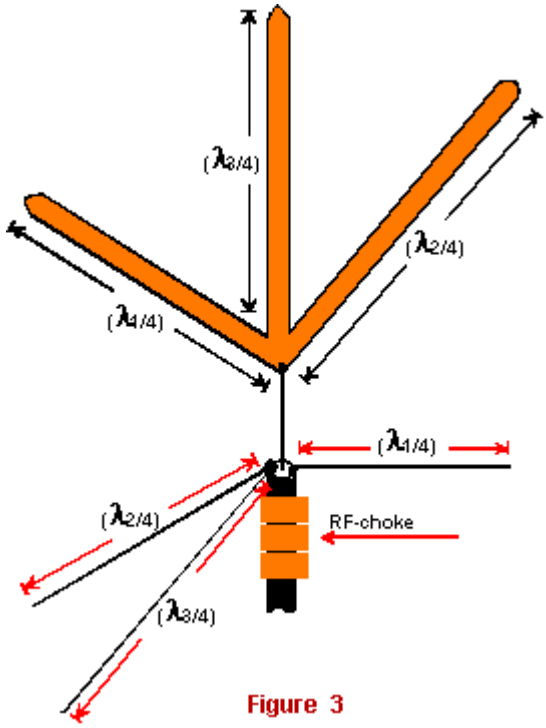


Figure 3

Figure 4 shows a simple design suitable for 6 - to 30-M. Vibrators are screwed to a strong metal angle.

Figure 4 A three range antenna on a metal angle

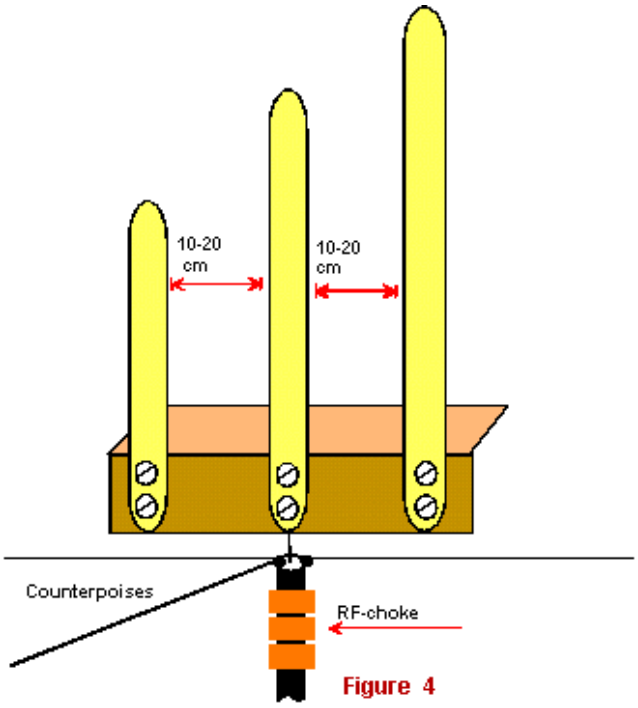


Figure 4

Distances between the vibrators are 10 -30 centimeters. It is decrease the mutual influence of the vibrators to each other.

Antenna Adjusting: The antenna is adjusted by changing lengths of the vibrators. It is not complicated. One way is to move vibrators relatively the metal base, as it is shown in **Figure 5**. Do it carefully, because the vibrators have mutual influence to each other. It needs to do additional holes on to end of the vibrators for realization of the way. It is possible to do one of the vibrators. This method always gives a good result.

Figure 5 A three range antenna adjusting

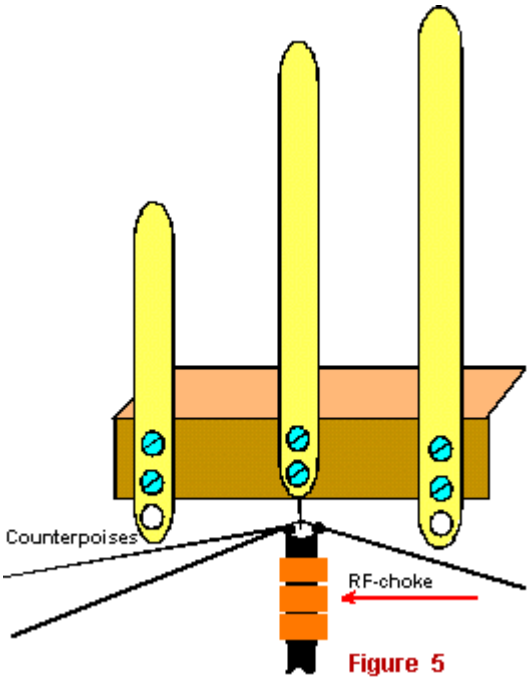


Figure 5

Other way is to change lengths of the upper ends of the vibrators. The vibrators ends made from thick copper or aluminum wire. The wire may be shortened, move in the side, as it is shown in **Figure 6**. But at the way an amateur must have access to ends of the antenna.

A three ranges antenna for the low ranges

Figure 7 shows a simple design suitable for 40 - to 160-M. Vibrators made from a copper wire in diameter 1 to 2 mm. Vibrators have length (λ/4)*1.1. Each vibrator is matched with coaxial cable with help of its own a 'shortening 'capacitor. The shortening capacitor can have 100-pF at ranges of 6- to 17-M, 150-pF at ranges of 20- and 30-M, 200-pF at ranges of 40-80 meters, 250-pF at 160-M. The shortening capacitors should be placed in a whether- proof box.

Figure 8 shows another simple design suitable for 40 - to 160-M. Vibrators made from a copper wire in diameter 1 to 2 mm.

Figure 6 A three range antenna tuning with the help of thick wire

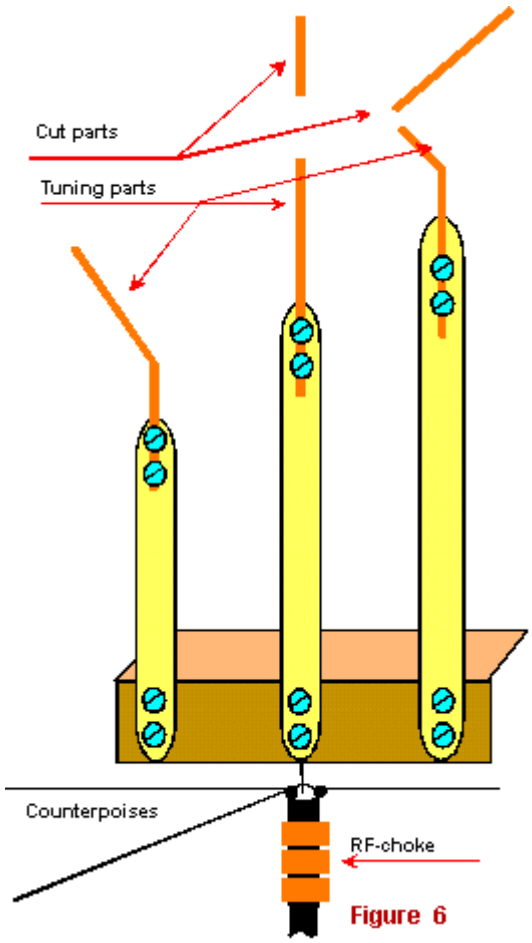


Figure 6

Vibrators have length (λ/4)*(0.5-0.9). Each vibrator is matched with coaxial cable with help of its own a 'lengthening' coil. You can use this design if you have a lack of place.

It is not wise to use more than three vibrators for a multi- range vertical antenna, because overall efficiency of the antenna drops in this case. Such multi- vibrators antenna will be too complicated at adjusting.

Remember: Two and more resonance (a quarter wave) counterpoises for each operation range of the antenna should be used. However, if the antenna is placed at a small altitude above a metal roof and the braid of the coaxial cable has a good electrical contact with the metal roof, the antenna could be used without any counterpoises.

RF – choke should be used: An RF- choke on the coaxial cable should be installed at feeding terminals. . The RF-choke precludes leaking of RF- currents on to outer braid of the coaxial cable. Without the RF-choke the outer braid of the coaxial cable serves as a radiating part of the vertical antenna. It gets to TV and RF- interferences when the antenna operates on transmission. 10 - 30 ferrite rings (permeability does not matter) hardly dressed on the coaxial cable end at the antenna terminal make the RF-choke.

Figure 7 A simple design suitable for 40 - to 160-M

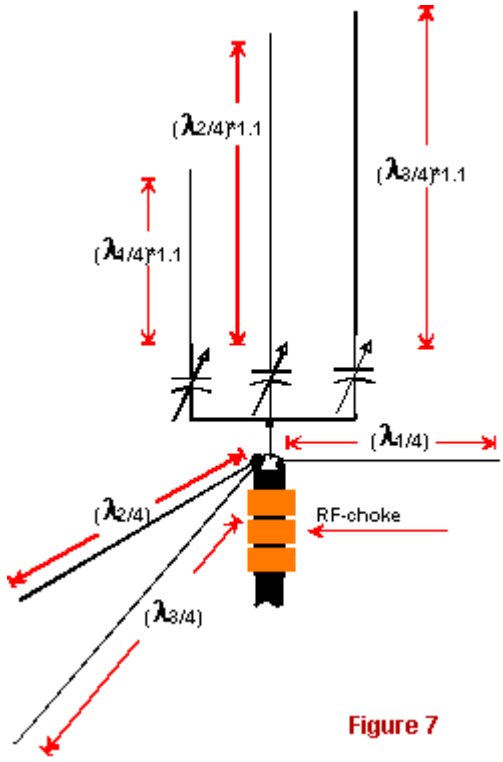


Figure 7

Figure 8 A simple design suitable for 40 - to 160-M with 'lengthening' coil

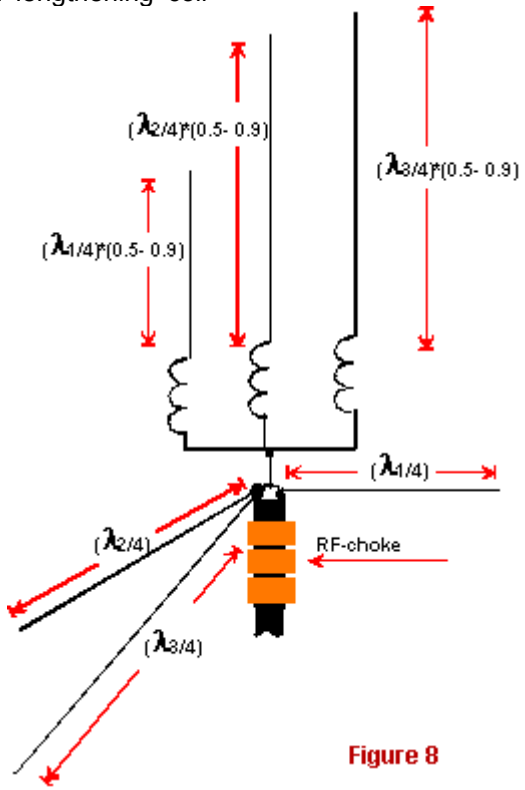
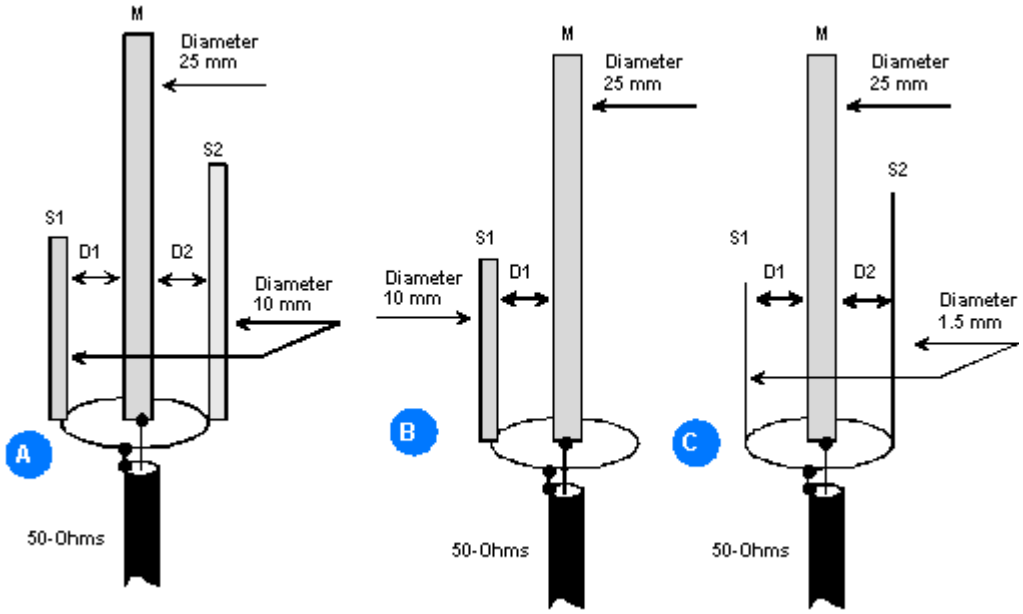


Figure 8

Practical Design of Open - Sleeve Antennas for Upper Amateur HF - Ranges

By Dmitry Fedorov, UA3AVR



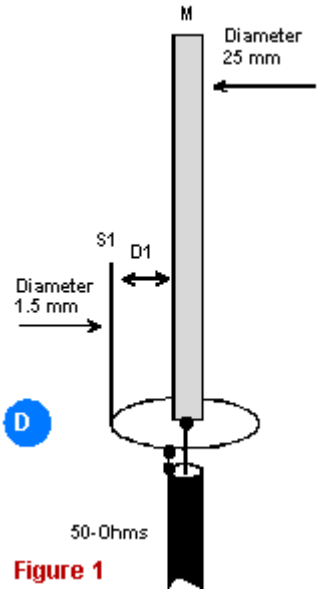


Figure 1

Open Sleeve antennas presented at this article were designed by UA3AVR (Reference: Dmitry Fedorov (UA3AVR).: Multi-range vertical Open Sleeve.- Radiomir. HF and VHF, 2001, #8, pp. 34-36). **Table 1** shows data for the Open Sleeve **Figure 1** shows the design of the antennas

Band, m	Length M, mm	Length S1, mm	Distance D1, mm	Length S2, mm	Distance D2, mm	**Figure 1**
20; 14; 10	5168	3407	220	2573	200	A
14; 10	3630	2527	220	-	-	B
20; 14; 10	5149	3451	220	2601	200	C
14; 10	3432	2567	210	-	-	D

Multi- Range Vertical Antenna UA1DZ

by Igor Grigorov, RK3ZK

Antenna history: *Antenna UA1DZ is a very interesting multi- range vertical antenna designed by known Russian radio amateur UA1DZ. The antenna was very popular in use in the former USSR. Russian radio amateurs widely use the antenna at present days also. The antenna works with a low SWR on 40-m, 20-m and 15m. Firstly UA1DZ told about his antenna in the ether, and after that, lots Russian radio amateurs have did the antenna and Antenna UA1DZ became very popularity. First printing papers about antenna UA1DZ appeared in reference* ***[1]****. This antenna has gain 3,67 dBi at 40-m, gain 4 dBi at 20-m, gain 7,6 dBi at 15m (reportedly to VA3TTT, reference* ***[2]****).*

Antenna construction: Figure 1 shows the construction and matching device of multi-range vertical antenna UA1DZ (based on reference **[1]**). The vibrator of the antenna has the length in 9.3 meters and four counterpoises of the antenna have length in 9.4 meters. Why has the antenna such sizes? Well, for his multi range antenna UA1DZ used an old military vertical antenna and this one had such sizes.

If you have not such old military vertical antenna, of course, it is possible to do home made vibrator and counterpoises! The vibrator and counterpoises must be made from copper or aluminum stuff. Do not use iron wire for HF antenna at all! Iron does not work properly in HF transmitting antennas, especially at upper amateur HF ranges.

Guys must be used with the antenna for providing wind strength. Use acryl cord or iron wire “broken” by insulators to one - meter lengths. Base insulator should have high mechanical and electrical strength because antenna vibrator has a large weight and there is high RF- voltage across the base insulator in transmitting period.

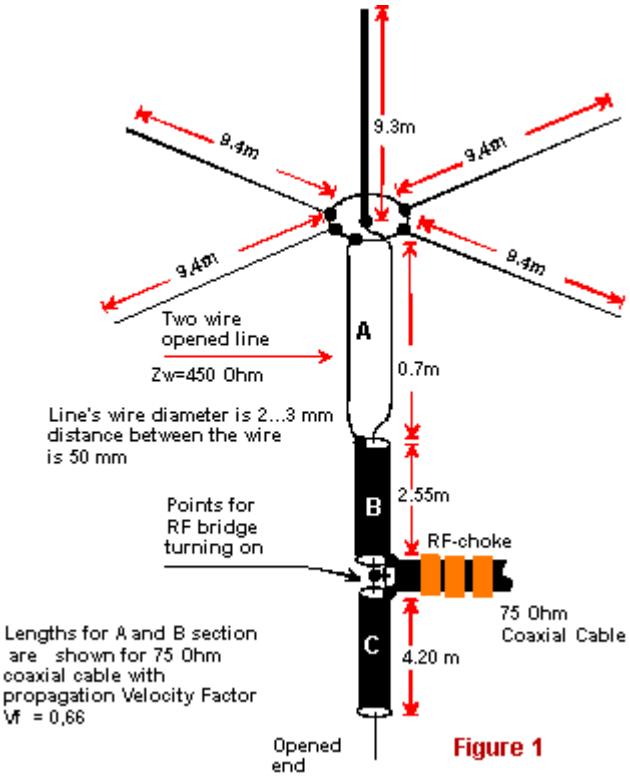


Figure 1

Matching device: It is made on one length of two – wire opened line and two lengths of a 75- Ohms coaxial cable. With the matching device the antenna can work on ranges 40-m, 20-m and 15m with a SWR in coaxial cable no more than 2:1. Two wire opened line “A” does initial matching the antenna input impedance with feeding coaxial cable. The line has characteristic impedance of 450 Ohm and one meter initial length. As usual, the line has ended length about 0.7- meter.

Coaxial cable “B” with characteristic impedance of 75-Ohm and with length 2.5 meters makes further matching for input impedance of the antenna system with feeding coaxial cable. An opened on the end length of coaxial cable “C” makes compensation of a reactive part of the input impedance of the antenna system.

Two wire line (part A) and the matching parts B and C must be placed not less the 50 centimeters above the roof. Parts A and B should be placed in straight line. It is possible to coil the part C in a bay.

Antenna tuning: The antenna UA1DZ is tuned as follow.

- An RF bridge is turned to input terminal of antenna matching device (see **Figure 1**).

- Shift antenna resonance frequencies in amateur 40- and 15-m bands by gradually diminishing the length of matching section A. Five centimeters truncation the length of matching section A does frequency shift up to 200 kHz on 21 MHz, and up to 60 kHz on 7 MHz.

It is quite possible to tune the length of matching section A so, that antenna UA1DZ will have the resonance frequencies inside ranges 21 and 7 MHz. If the antenna UA1DZ has resonances on these ranges (40- and 15-m), it will have a resonance frequency inside 20-m range.

Two-wire opened line: It is possible to use either commercial made two-wire opened line either homebrew one. Remind, that two-wire transmission line with aerial dielectric and 450 Ohm characteristic

impedance has relation between the diameter of its wires and the distance between these wires nearly 20 (see **Figure 2**).

RF – choke should be used: An RF-choke should be installed on the coaxial cable at the antenna terminal. This RF-choke precludes leaking of RF currents on the outer braid of the coaxial cable.

Without such RF-choke the outer braid of the coaxial cable will serve as a radiating part of the vertical antenna. It causes big level of RF interferences when the antenna works on transmission. 10 -30 ferrite rings, hardly dressed on the coaxial cable at the antenna terminal, make the most simple an RF-choke. The place for a RF choke is shown in **Figure 1**.

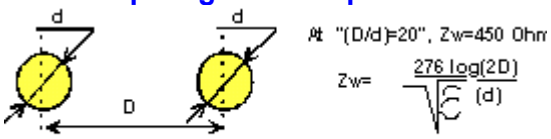


Figure 2

References:

1. RB5IM.: Ground plane UA1DZ. Bulletin UC 1993, C.27.

2. A. Barskiy, VA3TTT: About antenna l www.krasnodar.online.ru/hamradio

Hula- Hoop magnetic Loop

by ***Yuri Kazakevich***, EW6BN, , ew6bn@tut.by

After long QRT (birth of my daughter, changing my QTH) I was going again QRV!!!

So, I needed an antenna! But where can I install it? It was not possible to install any antenna on the roof of my house. I had only place for installation of an antenna, the place was my balcony of my house. Well, it was very place. What an antenna can install at the place? I though, it was only a Magnetic Loop Antenna.

I remembered, when I still went to school, I used a Magnetic Loop Antenna made from old coaxial cable for my work on CB - range 27 MHz. The antenna worked very well. Well, I decided to use a Magnetic Loop Antenna for my very restricted area for a work at 14 MHz.

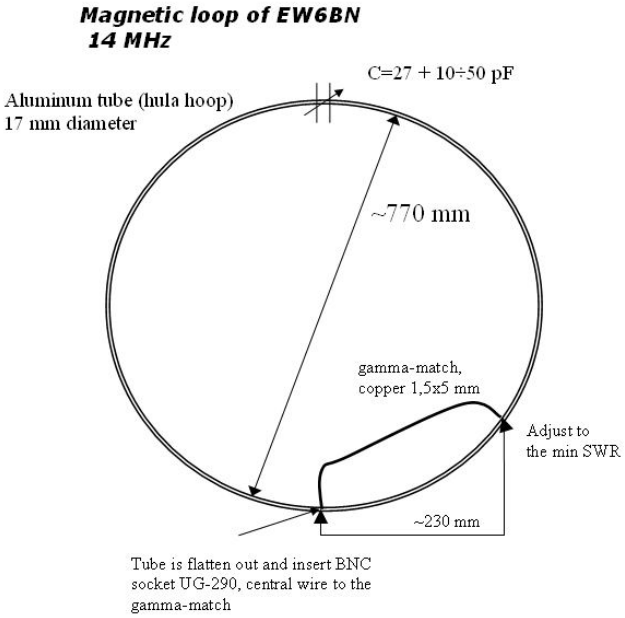
Lots information about Magnetic Loop Antennas I found in the Internet, in particular in reference **[1]**, it is a free e- book on antennas (in Russian).

I decided to make my Magnetic Loop Antenna on the basis of an aluminum hula - hoop. Hula – hoops in diameter of 77 centimeters and with 17 mm tube diameter were on sale in my local shop. The hula – hoop tuned at 14 MHz with two capacitor- one variable capacitor 10- 50 pF, and other, bridged to the variable capacitor, a fixed capacitor in 27 pF. The capacitors placed at the top the hoop. For my loop I used gamma feeding, because it has very high efficiency. **Figure 1** shows my Magnetic Loop Antenna. I have got 1:1.3 SWR with the gamma match.

The Magnetic Loop Antenna was installed on the third floor of a brick five-floor house. A wooden stick hold the antenna almost in one meter aside from the balcony. It was impossible to do a rotary design of the antenna for my conditions, so I just fixed the antenna on the line West – East. My house is situated at outskirts of the city, so, the West is opened, only one imperfection, a high-voltage power electric line on 110 kV is in 50 meters from my antenna….

On reception the antenna worked perfectly. But, unfortunately, there was a small handicap from the high-voltage power electric line.

Figure 1 Magnetic Loop Antenna



The antenna had very good results at transmission mode. See my first QSOs, that I have made straight away after installation of the antenna.

18:50 UTC, 13 July 2003:

I heard "CQ de G3KXV". I pressed on key – "G3KXV de EW6BN/QRP..."
And … "EW6BN/QRP de G3KXV" op Vic.
YES, the QSO is made!
I gave RST 579 QSB.
He gave me 569, also QSB, 100-w and a dipole, your mag loop 77 cm doing very well!

19:25, UTC, 13 July 2003:

HB9DRK/QRP stayed on CQ, he received my call, gave me 329, I gave him the info about my mag loop, and HB9DRK/QRP gave me a new rprt 559, he used 5-w and a delta.

Perfectly... My soul was singing, but I had to do QRT for a while…

So, my balcony Magnetic Loop Antenna allows me to be in the ether again and to do interesting QSOs over the World!

Reference:

1. Igor Grigorov: " Antennas for radio amateurs - 1998, Majkop, e-book, Available free at http: // cqham.ru/ftp/rk3zk/zip

EW6BN:A Field Operation

A HELICAL LOPP ANTENNA FOR THE 20 METERS BAND

By Vladimir Kuz'min, UA9JKW,
KuzminVI@pn.yungjsc.com

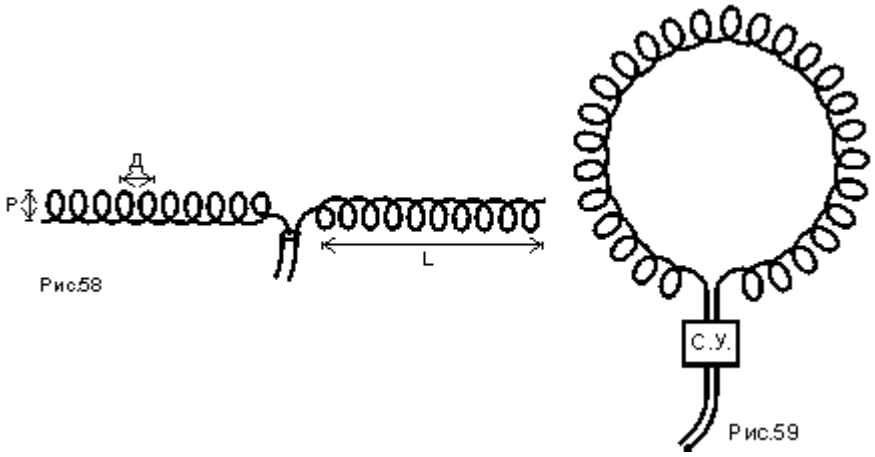
Two years back I have moved to Nefteyugansk (Russia, Siberia) where I could not receive the sanction to installation for a full-sized HF- antenna on the roof of my house. , So I began to do experimenters with short indoors antennas. Most success design of my indoor antenna is a design similar to ***Fig. 59****, given in* ***Reference 1****.*

I have used an inch OD plastic pipe to the form of the antennal. The pipe was bent in a hoop near1 meter diameter. Antenna has 580 turns (near 61 meters of length) of multicore isolated wire of 3 mm diameter with thickness of isolation of 1 mm. So, the spacing between turns is 2 mm. Antenna has SWR 1:1 to 50-Ohm coaxial cable to 14.100, bandwidth to SWR 1:1.5 is 300-kHz. I use a simple symmetrical device- 3 turns on a TV yoke ferrite core. Space from the antenna to the ceil is near 25 centimeters.

The antenna has quite good directed properties at rotation within 30-90 degrees the force of signals varied to 1-1,5 points on mine S-meter. I use a YAESU FT840 for my work in the ether. Change of polarization (at rotation of the antenna on the vertical side) appreciable changes has not given as well as change of feeding points has not given large change in the force of signals.

Helical Loop Antenna

Figure 58 & 59 from Reference 1



In the last summer I experimented and hung up of the antenna behind my balcony at 1.5 meters from a wall. I have received a significant improvement of the work of the antenna The antenna does very good operation in the ether, better than others indoors antennas. It gives low industrial noise and kills all TVI.

Reference:
I. Grigorov. Antennas for radio amateurs. - Majkop, 1998.

Get free the book from

http://www.cqham.ru/

UA9JKW at his shack

Top Load at Vertical Antennas

All amateurs know if at a vertical antenna a top load it is used, the self –resonance of the vertical antenna would be lower then a vertical without the top load. How a top load does influence to antenna resonance?

At **Reference 1** I found a very interesting table having the data. I have proved the table with **MMANA**, all okey, the table gives very reliable data, so it is possible to use it at many situations. Figure 1 shows different top loads. Data for loaded effect for the top load is shown in Table 1 given at **Reference 1**. K is coefficient: K = W/L, where W is a resonance wavelength for the vertical antenna, L is antenna length from the ground to the top load.

As it is seen, the “umbrella” top load (**Figure 1e**) gives the most effect on the resonance of a vertical antenna. For example, if to use an umbrella load for a vertical antenna in five meters height, the antenna quarter wave fundamental resonance wavelength would be changed from 20 to 50 meters!

Reference:

1. Polyakov V. Technique of radio: Simple AM receivers. – Moscow, DMK-Press, 2001.

73! I.G.

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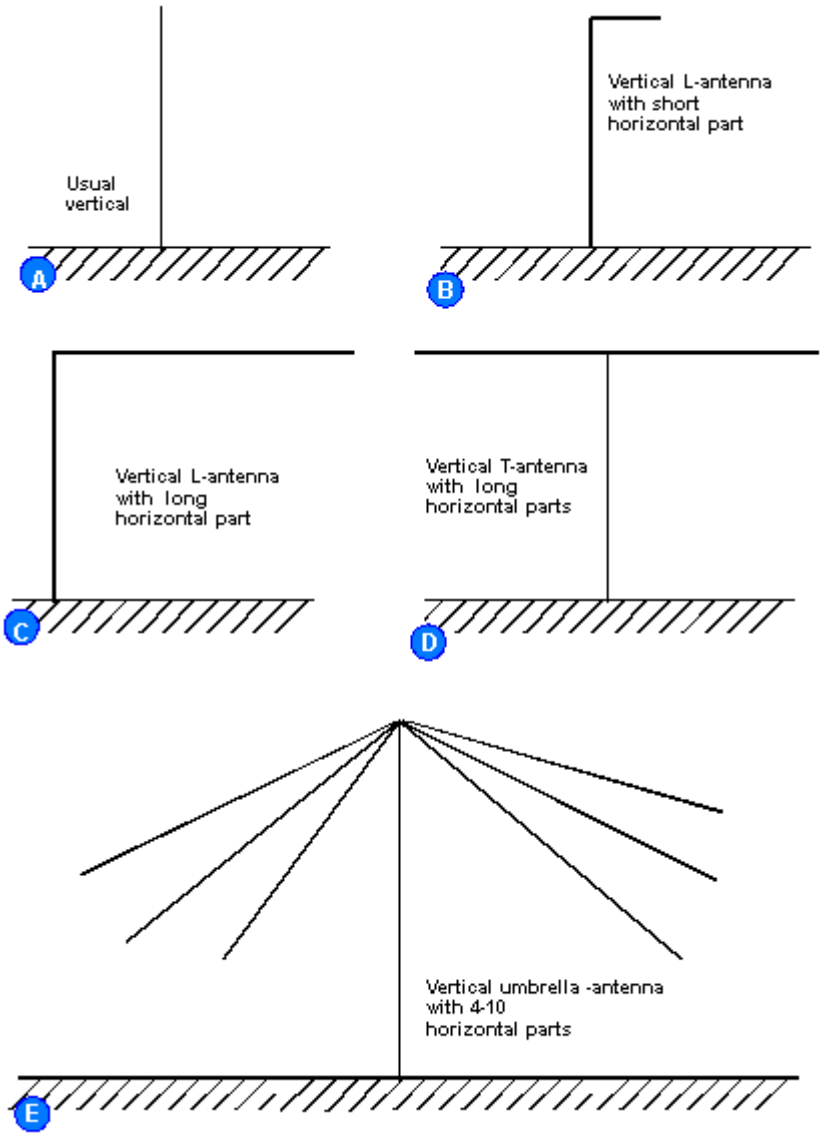

Figure 1

Kind of an antenna	K
Simple vertical **Figure 1a**	4
Inverted L with short horizontal part **Figure 1b**	4.5-5
Inverted L with long horizontal part **Figure 1c**	5-6
T- antenna with long horizontal part **Figure 1d**	6-8
Umbrella antenna with 4-8 wires **Figure 1e**	6-10

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THREE BAND UNIVERSAL RZ3AE ANTENNA

Evgeniy, RZ3AE

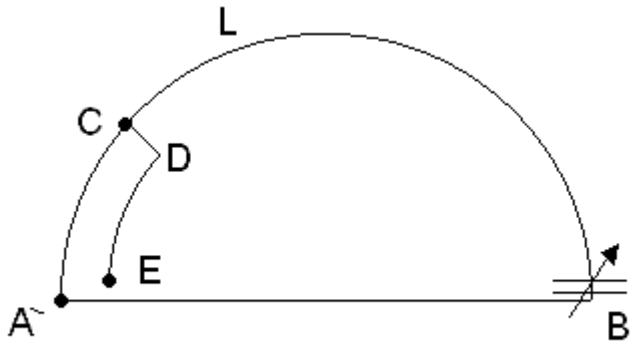
bort3@narod.ru

For several years I use to a simple and rather effective home – made antenna for a work from my balcony, a hotel window, from the ground in a radio- expedition and from my car. I with my friends have made a dozen such antennas and all the antennas work very well. One antenna, in depend of its dimension, works at three old amateur ranges- 10, 15,20 or 15, 20,40, or 20,40, 80.

Antenna description:

Figure 1 shows the antenna. Antenna wire is a tube or copper or bimetal rod of 5-12-mm diameter (#0000—5 AWG). D-E wire is thinner then a-c-b wire in 2-5 times. Antenna is tuned by air (a vacuum capacitor is better!) variable capacitor with air-gap in 2-mm. The capacity is 5- 750-pF.

Figure 1 RZ3AE Antenna



Antenna ratio:

L = 1.57AB

CD = 5-8 centimeters

AC ≈ 0.2L

Most high frequency for the antenna is: 4 (AB+L).

Antenna adjustment

To run a QRP power at most high frequency for the antenna. Move a crosspiece CD to find the minimum SWR. Check position of the crosspiece. To run a QRP power for next working band of the antenna. Move a crosspiece CD to find the minimum SWR. Check position of the crosspiece. And so on.

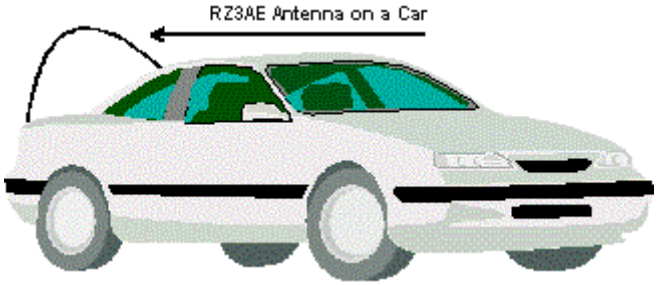
Antenna operation

To stand the crosspiece by manually or by RF-relays for choosing band. The antenna has very high directivity, so, choose needed position for the antenna. Switch your transceiver and enjoy!

Antenna results

I tried the antenna at different conditions and everywhere the antenna works well, from my house, from my car (the antenna is placed at boot of my car), from a field. I use to the antenna with IC-706 MK2G.

I wish all good luck! 73!



Fast Made a Half - Wave Antenna for 80 Meters

By Igor Grigorov, RK3ZK

G-QRP-C
6363

The antenna was made by me in one of the hot summer days near five years back. I was going for weekend to my bungalow and I decided to take my home- brew 80 – meters transceiver with myself. I had no antenna for the transceiver. So, I needed to do any antenna, but I had no time as no quality stuff for doing this one. I opened my box with old tips… and… Thirty minutes while I have had a new antenna that served me several years!

Figure 1 shows the all antenna system. I have done a half wave antenna with "bottle" matching device. As you can see a wire in long of 40 meters (a half wave antenna) is matched with 50-Ohm output of my transceiver with help of a parallel circuit ("bottle" matching device) – it is L1C1 in **Figure 1**. Spool L2 has not electrical connection with antenna circuit. RF energy is transferred from antenna to the spool only by magnetic field, that reduces the level of static interferences at receiving mode. The counterpoise has length of 20 meters of a naked copper wire in diameter of 1,5 millimeters (#14 AWG). I used a wire from an old burned down electrical transformer 220-V/12-V. The counterpoise serves as electrotechnical both as radio ground for the antenna. At operation time of the antenna the counterpoise is placed on the ground in any position (straight or bending). To short static electrical charge from antenna wire to ground is main task of the counterpoise. Not wise to use a long antenna in field without an electrotechnical ground, because in the first it is unsafe, and in the second, the antenna is very rustle on reception without an electrotechnical ground.

Figure 2 shows the construction of the matching device. I used a half - liter plastic bottle in diameter 60 millimeters from mineral water. C1 is attached at a side of the bottle with help of a strong copper wire in diameter of 1 millimeter (#18 AWG). L1 has 15 turns of copper wire in diameter of 1,5 millimeters (# 14 AWG), length of winding is 70 millimeters.

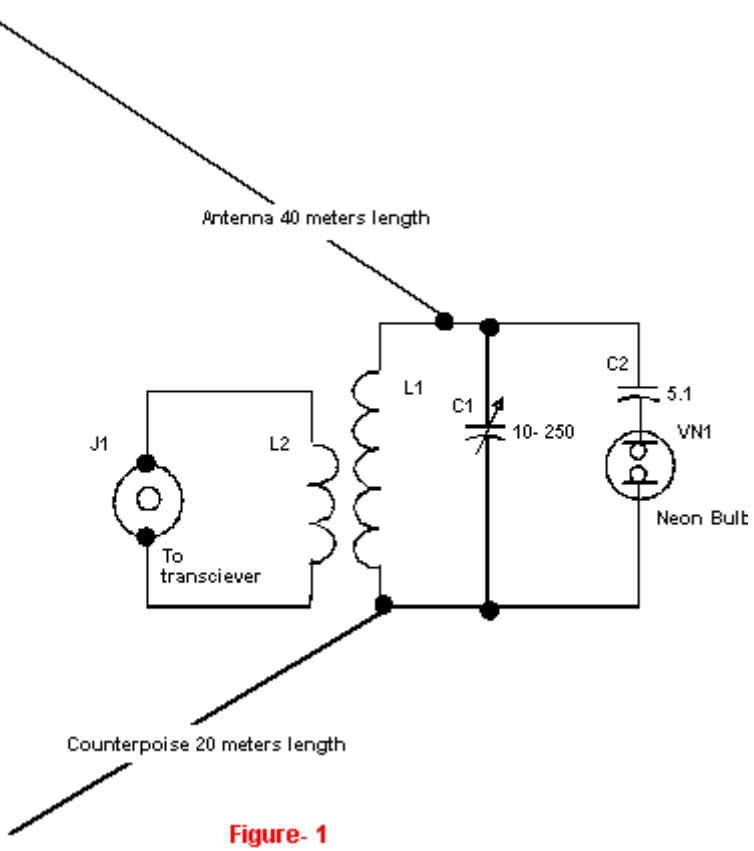


Figure- 1

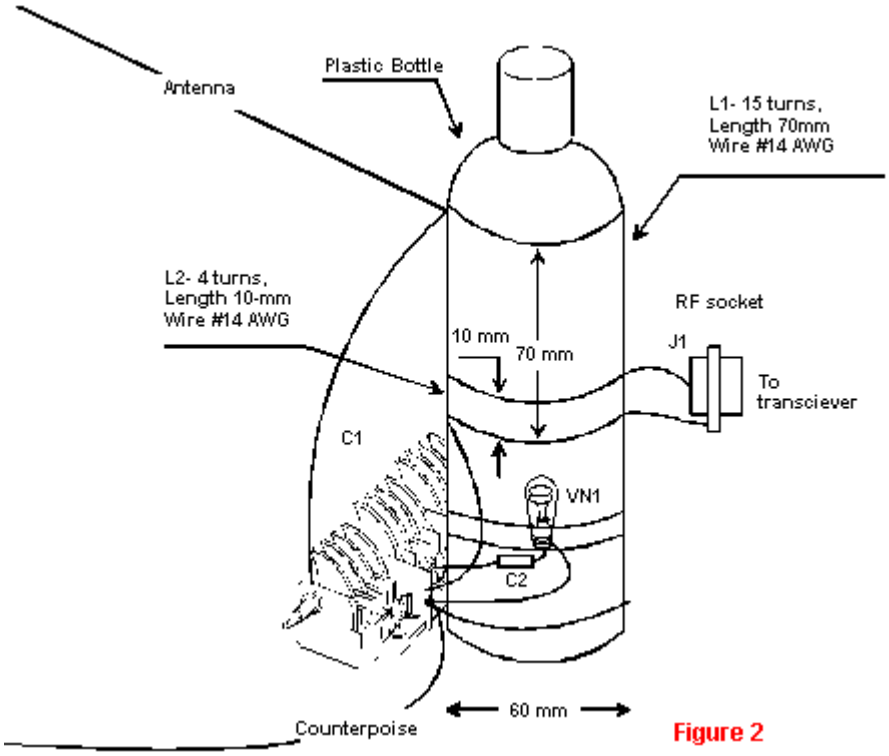


Figure 2

L2 is placed at the bottom of L1. L2 contains 4 turns of copper wire in diameter of 1,5 millimeters (# 14 AWG), length of winding is 10 millimeters. Ends of L2 are directly soldered to J1 RF – socket. VN1 is attached by a piece of Scotch to the bottle. Antenna is tuned by max glow of VN1.

The antenna works very effectively when the upper end of the antenna at lengths of five or more meters above the ground. I don't use an end antenna insulator. A long synthetic rope can simply be attached to the upper end of the antenna. The down end of the antenna could be just near the ground. A coaxial cable having any reasonable length can be between "bottle" ATU and a ham transceiver. **Figure 3** shows the antenna at field operation. Of course, it is very possible to use the antenna for stationary work from a ham shack.

The antenna works very well, and I recommend try it!

73/72!

One more a website devoted QRP!

US1RCH Page
www.qsl.net/us1rch

http://www.qsl.net/us1rch/

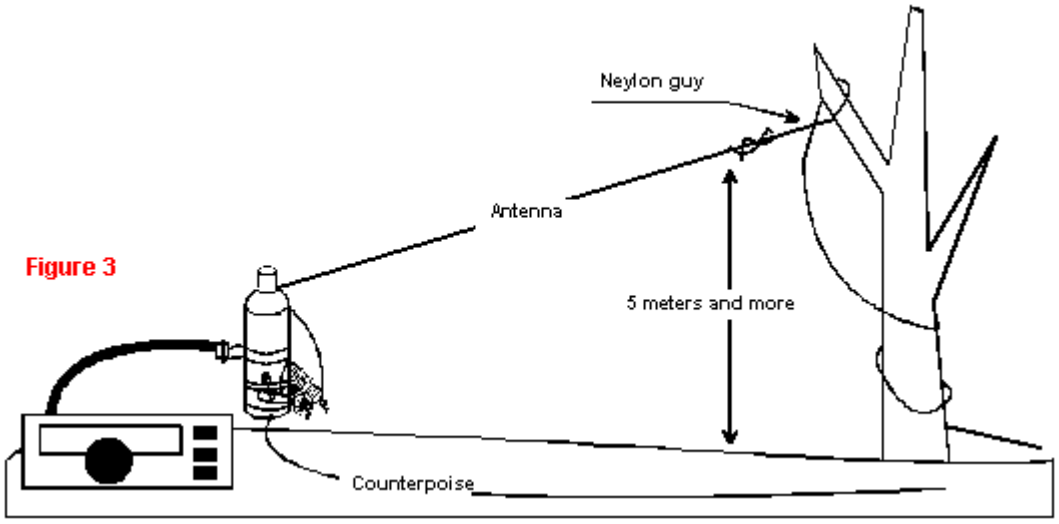


Figure 3

WINDOW DIPOLE ANTENNAS WITH CAPACITIVE LOADS FOR THE 6 AND 10 METERS BANDS

by Igor Grigorov, VA3ZNW
mailto: antentop@antentop.org

It is possible to install a dipole antenna with capacitive loads for the 6 and 10 meters bands at a standard window with sizes 140x150 or 140x210 centimeters. The design of that window antenna for the bands can be simple as well as the antenna impedance can be easy matched with 50-Ohm coaxial cable. That dipole antenna with capacitive loads installed at upper floor of a high-rise building can provide DX- QSOs.

However, a dipole window antenna with capacitive loads for bands low the 10-meters, if this one is installed at a standard window with sizes 140x150 or 140x210 centimeters, has low input impedance and narrow bandwidth so the antenna is hard to match. Hence antennas for bands low the 10 meters are not discussed in this article.

Types of a Dipole Antenna with Capacitive Loads

There is several ways to install a dipole antenna with capacitive loads at a window. The best way is to install a dipole antenna with capacitive loads by the center of the window. In that case the antenna can be installed at any house as made from a brick or wood as well as made from a concrete. Let's name the antenna "antenna central installation." If a house made of a brick or wood it is possible to install the dipole antenna with capacitive loads by up or down of the window. Let's name the antenna "antenna up or down installation." Of course, a non metal window-frame works better the metal one.

Feeding Coaxial Cable of a Dipole Antenna with Capacitive Loads

Ferrite rings (5- 20 ring with any permeability) installed at two ends of the coaxial cable going from TX to the antenna prevent RF- currents going from the antenna to TX. Since the rings do balun's job. Fasten the rings at the coax with a Scotch. The coaxial cable going from the antenna to the window-sill should be placed athwart to the antenna. However the coaxial can be placed as you want at you room.

Stuff for a Dipole Antenna with Capacitive Loads

A dipole antenna with capacitive loads can be made of a flexible multi- wire cable as well as of a strand wire. Any wire is good as naked as well covered by plastic isolation. Diameter of the wire can be near 1- 2 millimeters (12- 18 AWG). Use wire as much thick as possible. Compare to antenna made from thin wire thick antenna has wider bandwidth. It is wise (because it is cheap) to do a dipole antenna with capacitive loads without ends insulators. The antenna can be installed with help of a rope or plastic (as well as fishing) cord. A dipole antenna with capacitive loads of up or down installation can be installed directly (with help of nail or staple) at plastic or wooden window frame.

Window Dipole Antennas with Capacitive Loads for 6-meters Band

Figure 1 shows a schematic (**Figure 1a**) and design (**Figure 1b**) of a window dipole antenna with capacitive loads of central installation. **Figure 2** shows a schematic of a window dipole antenna with capacitive loads of up or down installation. The design of the antenna is similar to design shown at **Figure 1b**. The design of the both antennas is simple. Two ropes are installed at two ends of the window. Capacitive loads fastened to the ropes by thin wires or ropes. Third rope is installed at the center of the window. Antenna central insulator (made from a piece of any plastic or PC board) is fastened to the rope.

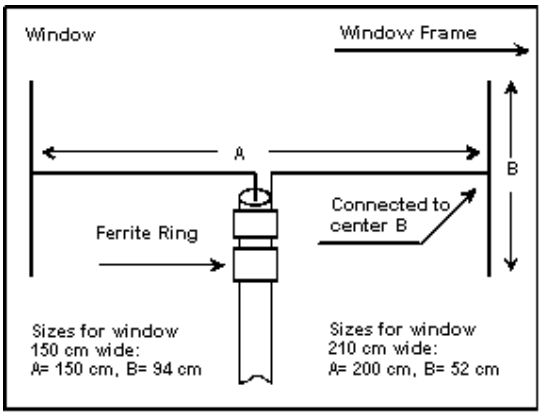


Figure 1 A window dipole antenna with capacitive loads of central installation

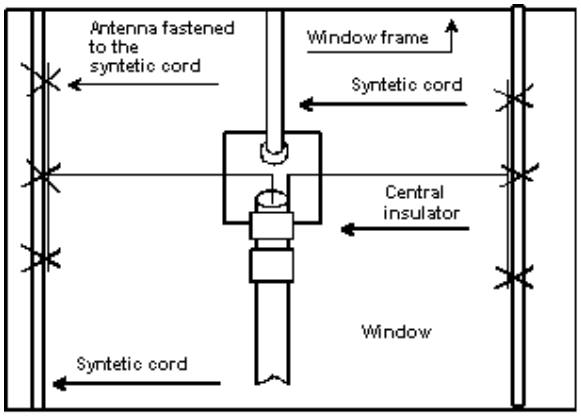


Figure 1B

Figure 1 A window dipole antenna with capacitive loads of central installation

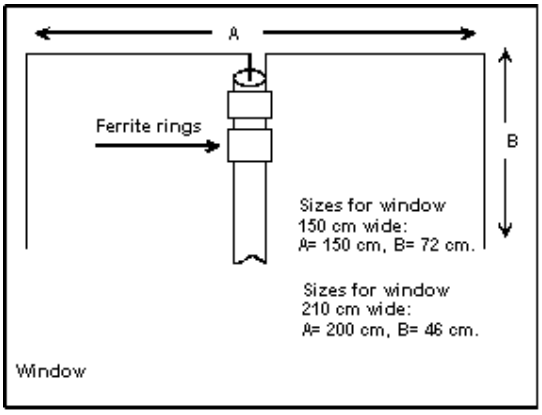


Figure 2

Figure 2 A window dipole antenna with capacitive loads of up or down installation

Adjustment of the both antennas is simple. A SWR-meter or HF- bridge (see References **[1]**) is connected to feed points of the tuned antenna. Gradually shorten 'moustaches' (symmetrically each moustache) of the antenna to minimum SWR or when antenna input impedance is active (has no reactive component) at needed frequency. At shortening moustaches the moustache wires roll up to a little coil.

Parameters of the Window Dipole Antenna with Capacitive Loads of Central Installation

Theoretical parameters of the antennas (copper, wire in 1-mm (18- AWG) diameter) were simulated with help of MMANA (see References **[2]**).

Figure 3 shows the input impedance of the antenna installed at window 150-cm wide. **Figure 4** shows the input impedance of the antenna installed at window 210-cm wide. Theoretical input impedance for 'narrow' antenna is 42-Ohms, for 'wide' antenna is 60- Ohms. The data are very good matched with my practical measurement of the antennas. A 50- Ohm coaxial cable should be used for feeding of the antennas. This one can be connected directly to antenna feed points, as it is shown at **Figure 1**. A 75-Ohm coaxial cable is possible to use for the antenna installed at wide (210 cm) window. **Figure 5** shows a SWR at 50- Ohm coaxial for 'narrow' antenna shown at **Figure 1**. **Figure 6** shows a SWR at 50- Ohm coaxial for 'wide' antenna shown at **Figure 1**. Theoretical gain for the antennas is near 1,5- 1,7 dBi.

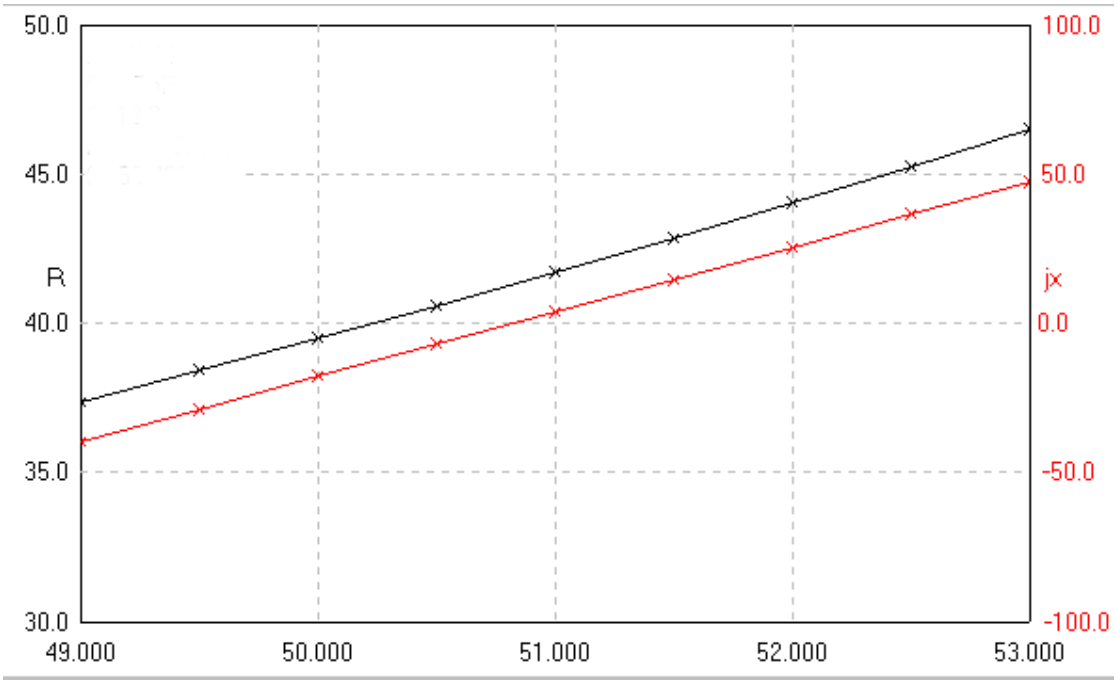


Figure 3 Input impedance of 'narrow' antenna

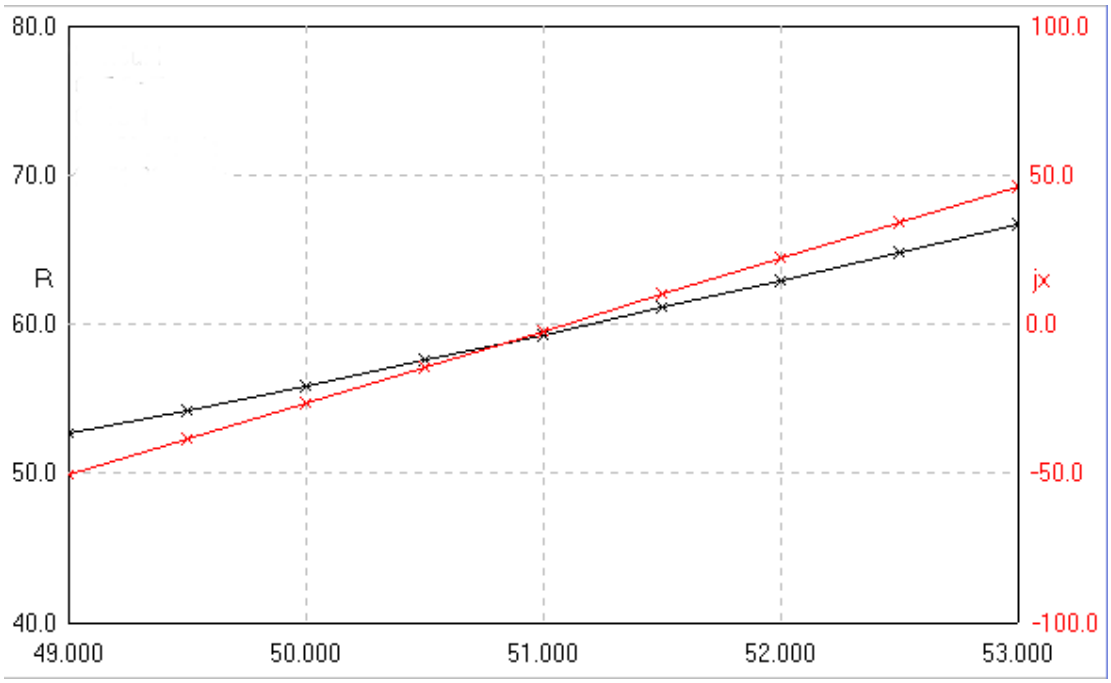


Figure 4 Input impedance of 'wide' antenna

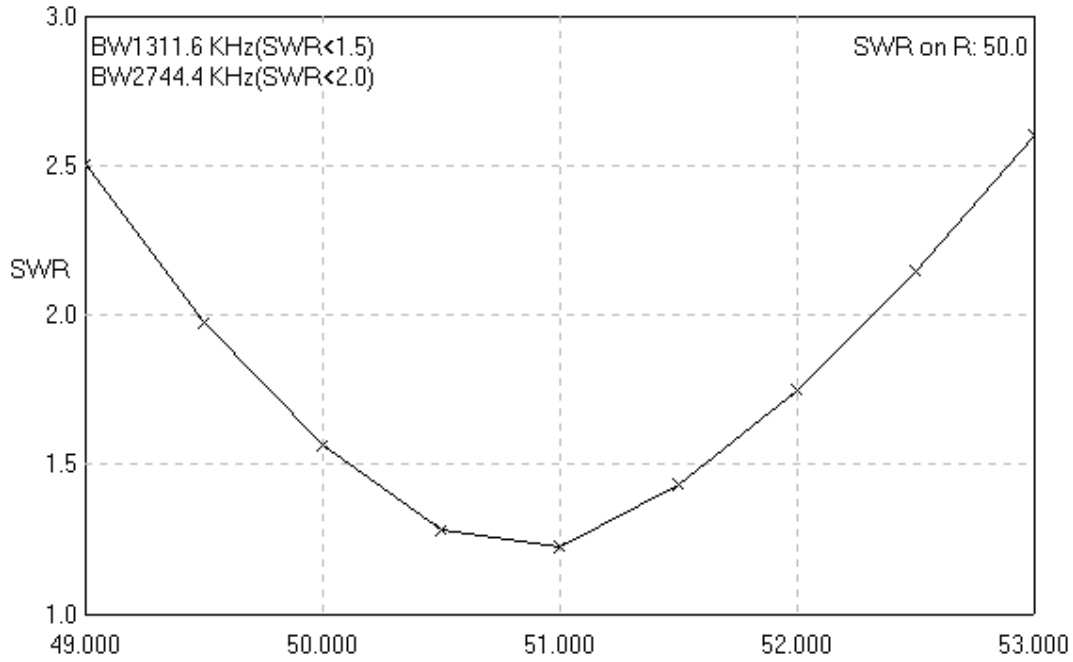


Figure 5 SWR at 50- Ohm coaxial for 'narrow' antenna

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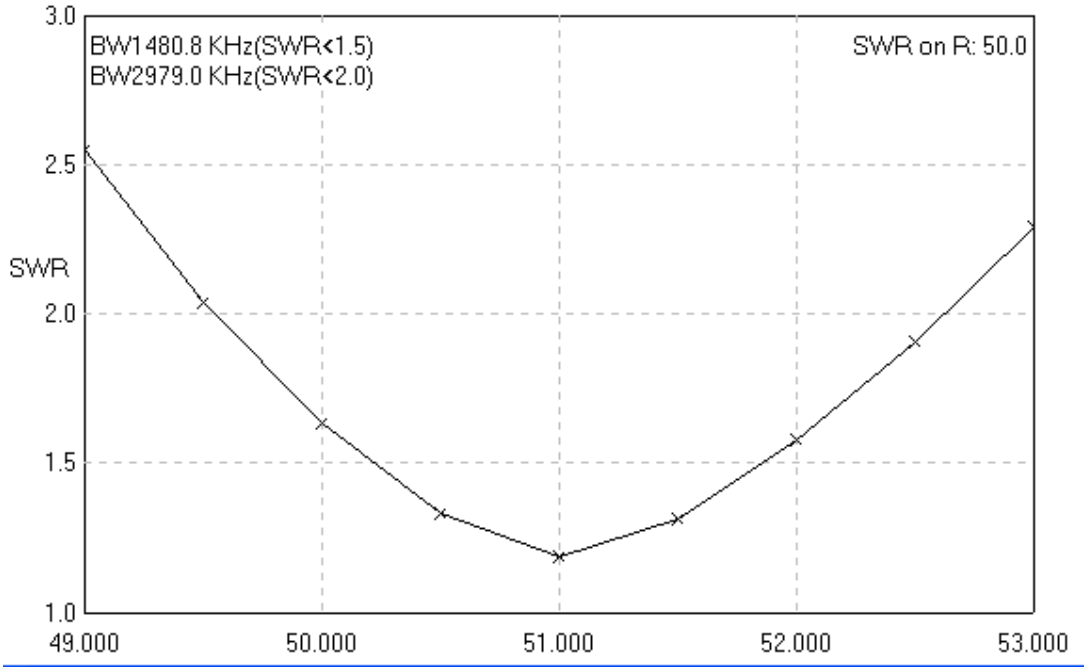


Figure 6 SWR at 50- Ohm coaxial for 'wide' antenna

A 'narrow' antenna of central installation has theoretical pass band 1300 kHz at SWR 1,5:1 at 50- Ohm coaxial cable, and pass band 2744 kHz at SWR 2:1 at 50- Ohm coaxial cable. A 'wide' antenna of central installation has theoretical pass band 1480 kHz at SWR 1,5:1 at 50- Ohm coaxial cable, and pass band 2979 kHz at SWR 2:1 at 50- Ohm coaxial cable. It is quite enough for working at 6- meters band especially since the real antenna has pass band wider the theoretical.

Parameters of the 6- meters Band Window Dipole Antenna with Capacitive Loads of Up and Bottom Installation

Theoretical parameters of the antennas (copper, wire in 1-mm (18- AWG) diameter) were simulated with the help of MMANA. **Figure 7** shows input impedance of the antenna installed at window 150 cm wide. **Figure 8** shows input impedance of the antenna installed at window 210 cm wide. Theoretical input impedance for 'narrow' antenna is 43- Ohms, for 'wide' antenna- 60- Ohms. The data are very good matched with my practical measurement of the antennas.

A 50- Ohm coaxial cable should be used for feeding of the antennas. This one can be connected directly to antenna feed points, as it is shown at **Figure 2**. A 75- Ohm coaxial cable is possible to use for an antenna installed at wide (210 cm) window.

Figure 9 shows a SWR at 50- Ohm coaxial for 'narrow' antenna shown at **Figure 2**. **Figure 10** shows a SWR at 50- Ohm coaxial for 'wide' antenna shown at **Figure 2**. Theoretical gain for the antennas is near 1,5- 1,7 dBi.

A 'narrow' antenna with capacitive loads of up or down installation has theoretical pass band 1377 kHz at SWR 1,5:1 at 50- Ohm coaxial cable, and pass band 2697 kHz at SWR 2:1 at 50- Ohm coaxial cable. A 'wide' antenna with capacitive loads of central installation has theoretical pass band 1393 kHz at SWR 1,5:1 at 50- Ohm coaxial cable, and pass band 2876 kHz at SWR 2:1 at 50- Ohm coaxial cable. It is quite enough for working at 6- meters band especially since the real antenna has pass band wider the theoretical.

Window Dipole Antennas with Capacitive Loads for 10-meters Band

Figure 11 shows schematic a window dipole antenna with capacitive loads of central installation. **Figure 12** shows schematic a window dipole antenna with capacitive loads of up or down installation. Antenna central installation can be installed at window 210-cm wide. Antenna up or down installation can be installed at window 150 or 210-cm wide.

The design of the both antennas is similar to design shown at **Figure 1b**. Two ropes are installed at two ends of the window. Capacitive loads fastened to the ropes by thin wires or ropes. Diagonal capacitive loads are spread by thin ropes. Third rope is installed at the center of the window. Antenna central insulator (made from a piece of any plastic or PC board) is fastened to the rope.

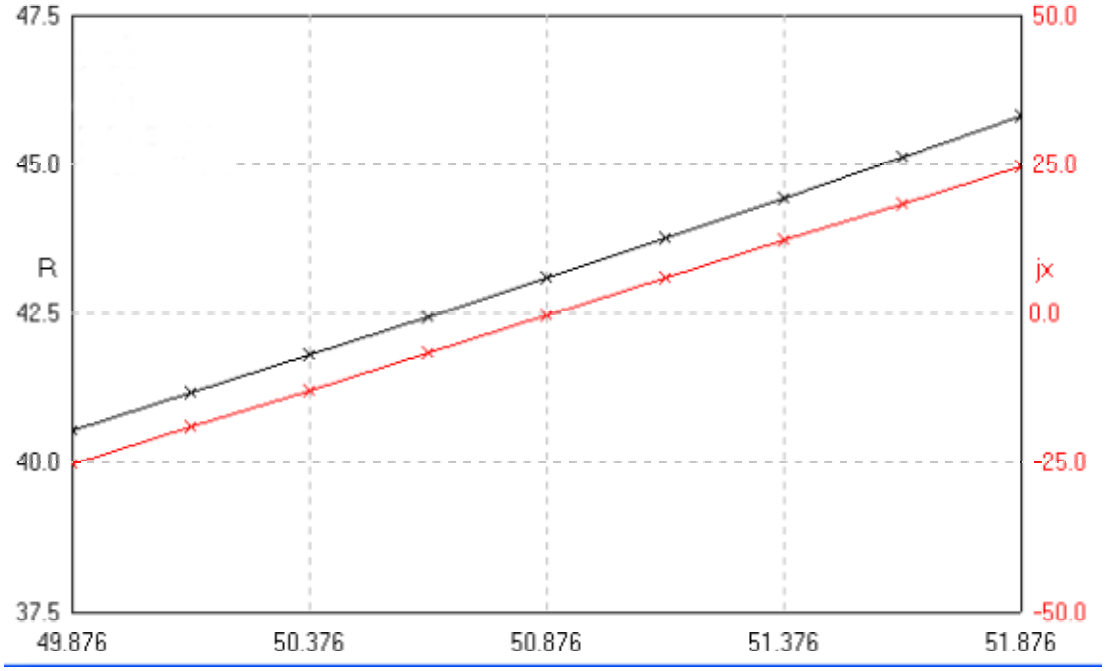


Figure 7 Input impedance of 'narrow' antenna

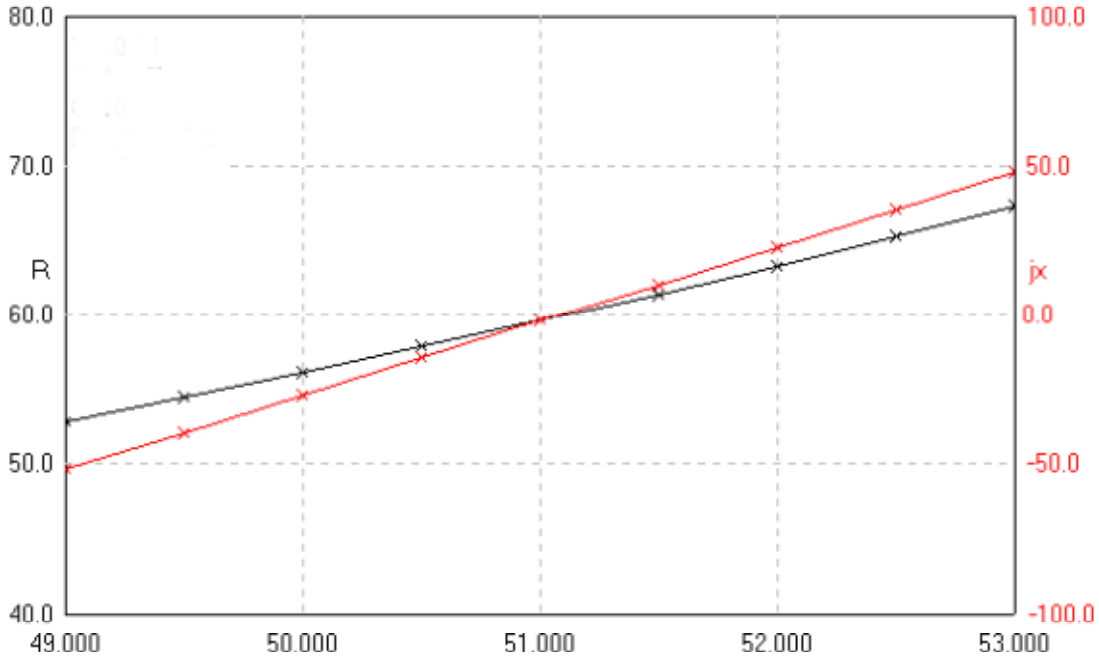


Figure 8 Input impedance of 'wide' antenna

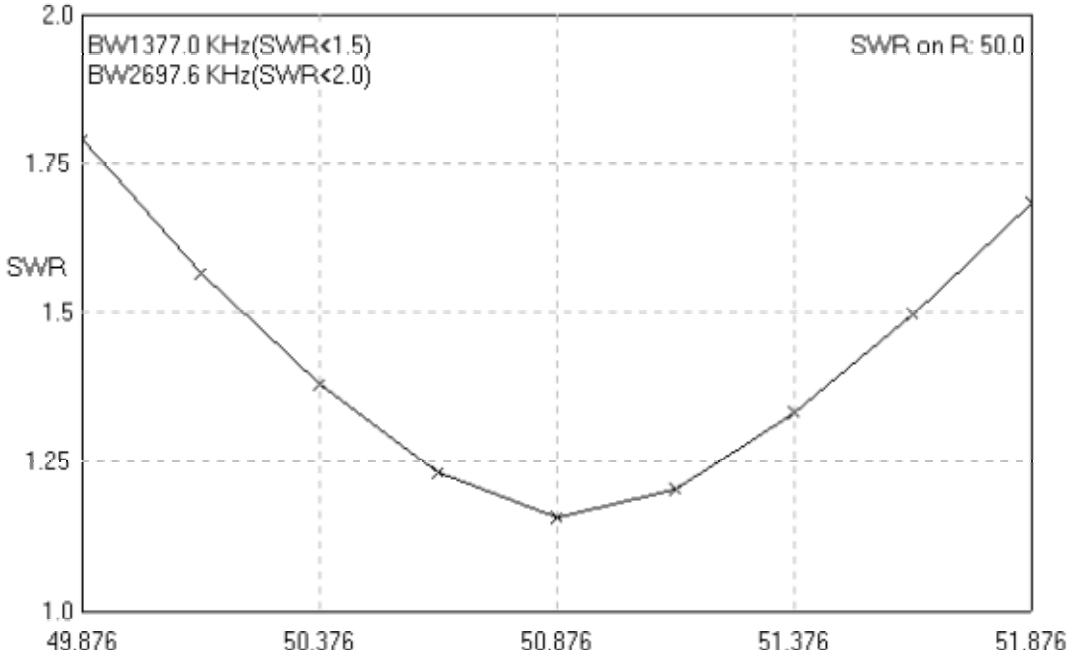


Figure 9 SWR at 50- Ohm coaxial for 'narrow' antenna

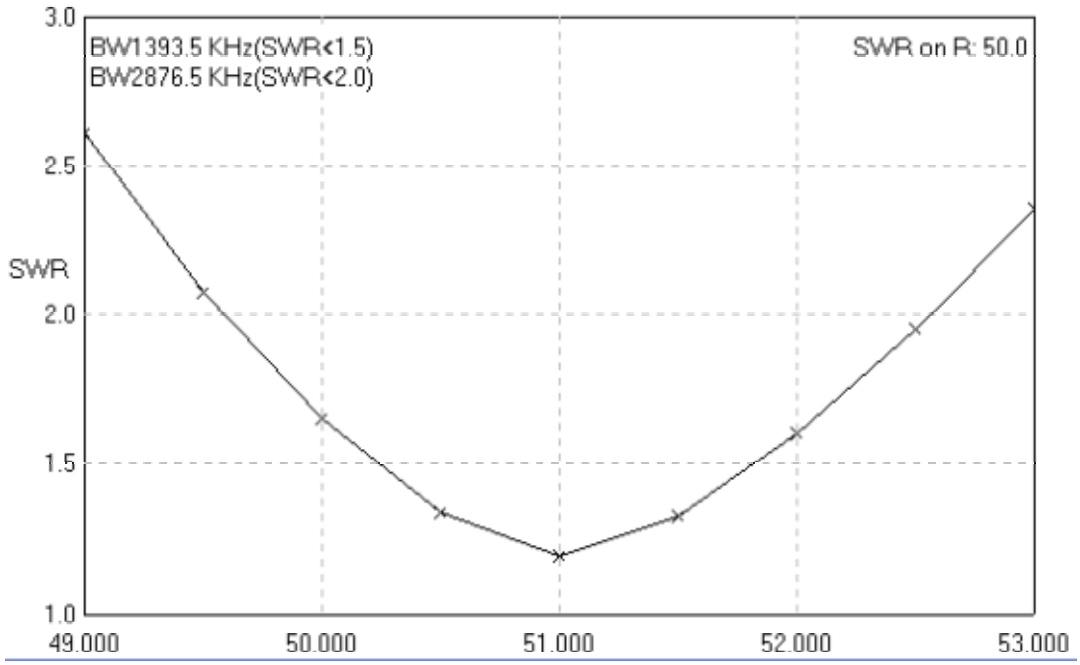


Figure 10 SWR at 50- Ohm coaxial for 'wide' antenna

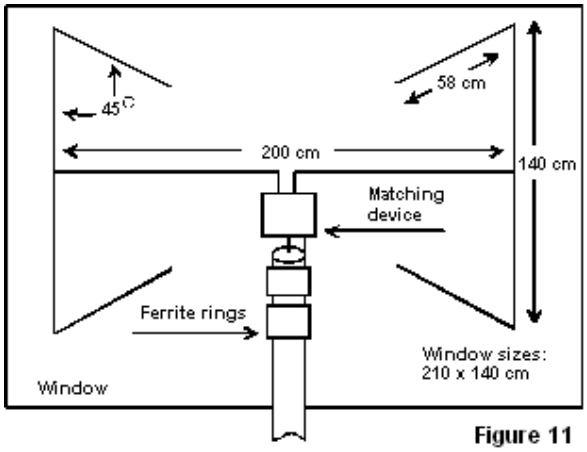


Figure 11

Figure 11 A window dipole antenna with capacitive loads of central installation

Adjustment of the both antennas is simple. A SWR-meter or HF- bridge (see References **[1]**) is connected to feed points of the tuned antenna. Gradually shorten moustaches (symmetrically each moustache) of the antenna to minimum SWR or when antenna input impedance is just active (have no reactance) at needed frequency. At shortening moustaches the moustache wires roll up to a little coil.

Input Impedance of 10-meters Band Window Dipole Antennas with Capacitive Loads

Theoretical parameters of the antennas (copper, wire in 1-mm (18- AWG) diameter) were simulated with the help of MMANA. **Figure 13** shows input impedance of the antenna shown in **Figure 11**. Theoretical input impedance of the antenna is 22- Ohms. Practically measured input impedance of the antenna was 30- Ohms. Losses in neighbor objects add the 8 Ohms. Theoretical gain for the antennas is near 1,5- 1,7 dBi.

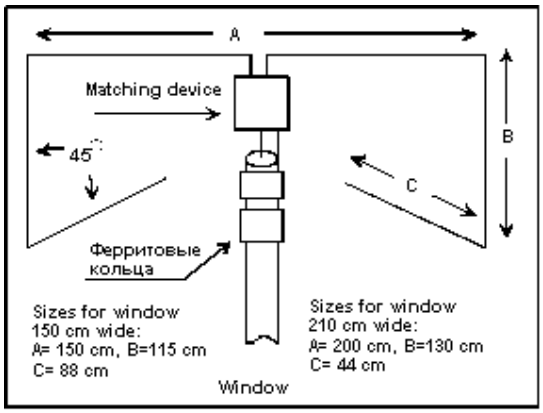


Figure 12

Figure 12 A window dipole antenna with capacitive loads of up or down installation

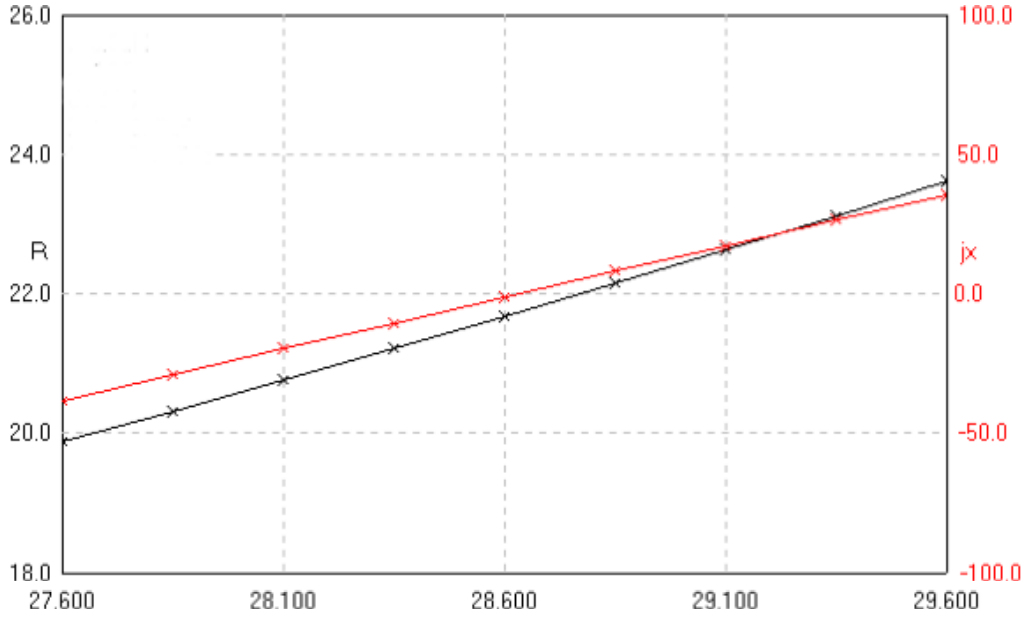


Figure 13 Input impedance of the antenna shown in **Figure 11**

Feeding of 10-meters Band Window Dipole Antennas with Capacitive Loads

Since 10- meters band window dipole antenna with capacitive loads has low input impedance a matching device must be installed between the antenna and feeding coaxial cable.

MMANA allows to simulate such matching device. **Figure 14** shows schematic of that matching device as well as data for different antennas. Of course, it needs adjust a little the L and C to particular antenna.

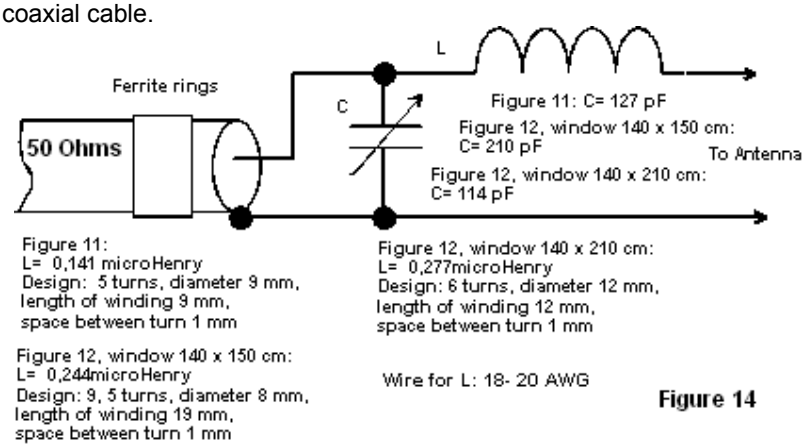


Figure 14 Matching device for antenna with capacitive loads

SWR of 10 – meters Band Window Dipole Antennas with Capacitive Loads

Figure 15 shows a SWR at 50- Ohm coaxial connected through a matching device (see **Figure 14**) to antenna shown at **Figure 11**. SWR was simulated by MMANA.

The antenna has theoretical pass band 375 kHz at SWR 1,5:1 at 50- Ohm coaxial cable, and pass band 750 kHz at SWR 2:1 at 50- Ohm coaxial cable. It is not enough for working at all 10- meters band. However, due the losses at neighbor subjects the pass band of the antenna is wider the theoretical one.

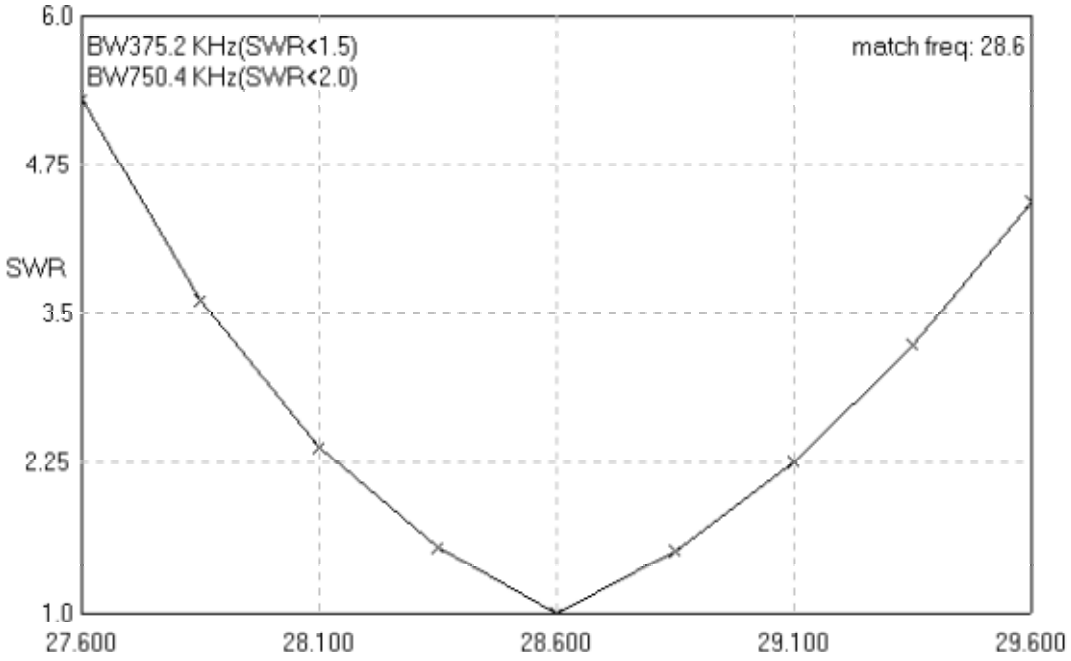


Figure 15 SWR at 50- Ohm coaxial connected through a matching device (see **Figure 14**) to antenna shown at **Figure 11**

Parameters of the 10- meters Band Window Dipole Antenna with Capacitive Loads of Up and Bottom Installation

Theoretical parameters of the antennas (copper, wire in 1-mm (18- AWG) diameter) (see **Figure 12**) were simulated with help of MMANA. **Figure 16** shows input impedance of the antenna installed at window 150 cm wide.

Figure 17 shows input impedance of the antenna installed at window 210 cm wide. Theoretical input impedance for 'narrow' antenna is 12- Ohms, for 'wide' antenna- 26- Ohms. Practically measured impedance is higher on 8- 10 Ohms the theoretical due losses at neighbor subjects. **Figure 18** shows a SWR at 50- Ohm coaxial connected through matching device (see **Figure 14**) to 'narrow' antenna (see **Figure 12**).

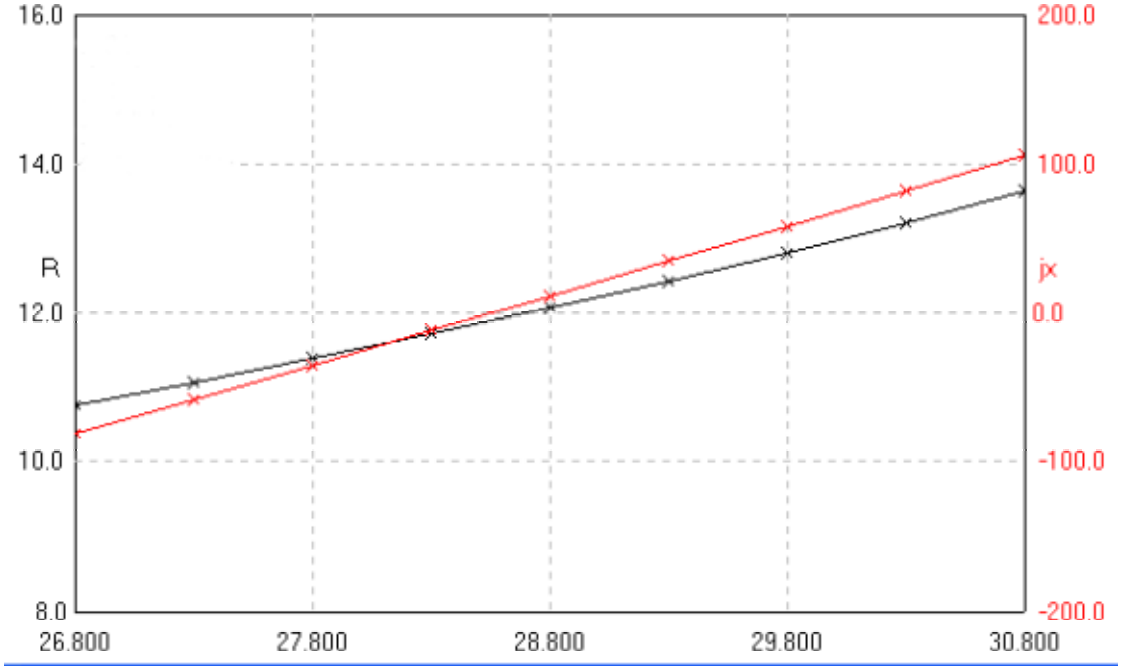


Figure 16 Input impedance of the antenna installed at window 150-cm wide

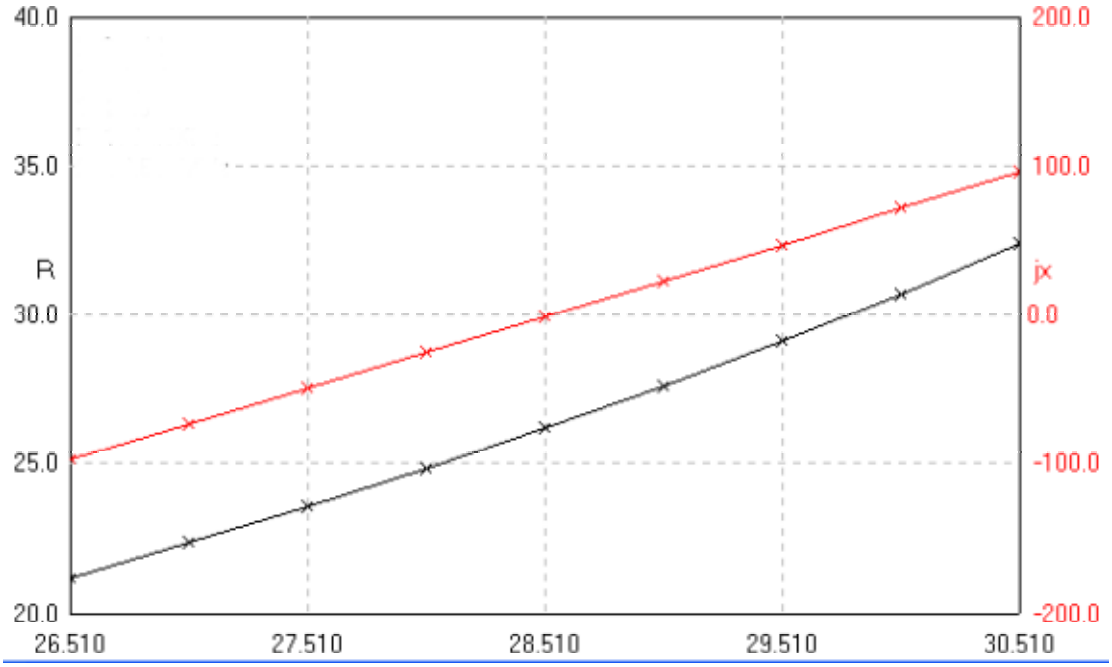


Figure 17 Input impedance of the antenna installed at window 210-cm wide

Figure 19 shows a SWR at 50- Ohm coaxial connected through matching device (see **Figure 14**) to 'wide' antenna (see **Figure 12**). SWR was simulated by MMANA. Theoretical gain for the antennas is near 1,1-1,3 dBi

A 'narrow' antenna with capacitive loads of up or down installation has theoretical pass band 157 kHz at SWR 1,5:1 at 50- Ohm coaxial cable, and pass band 314 kHz at SWR 2:1 at 50- Ohm coaxial cable. A 'wide' antenna has theoretical pass band 425 kHz at SWR 1,5:1 at 50-Ohm coaxial cable, and pass band 733 kHz at SWR 2:1 at 50- Ohm coaxial cable.

So those antennas can work only at a part of the 10 meters band. For working at all 10- meters band matching device can be retuned for needed frequency. However, the 'wide' antenna often works at all 10 meters band without retuning the matching device due losses at neighbor subjects.

References:

1. Igor Grigorov. Antennas. Matching and Adjustment. – Moscow. RadioSoft, 2002. ISBN 5-93037- 087-7

2. http://dl2kq.de/

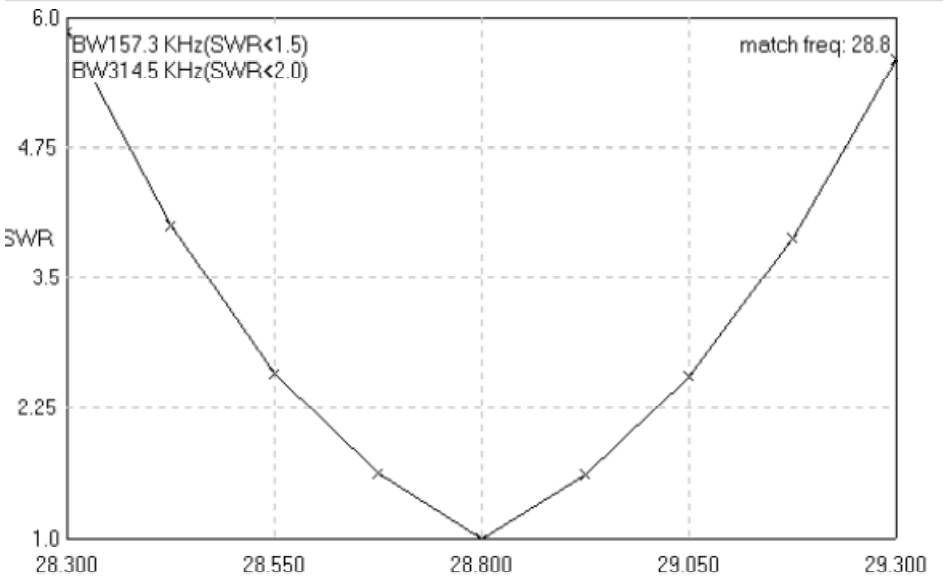


Figure 18 SWR at 50- Ohm coaxial connected through matching device to 'narrow' antenna

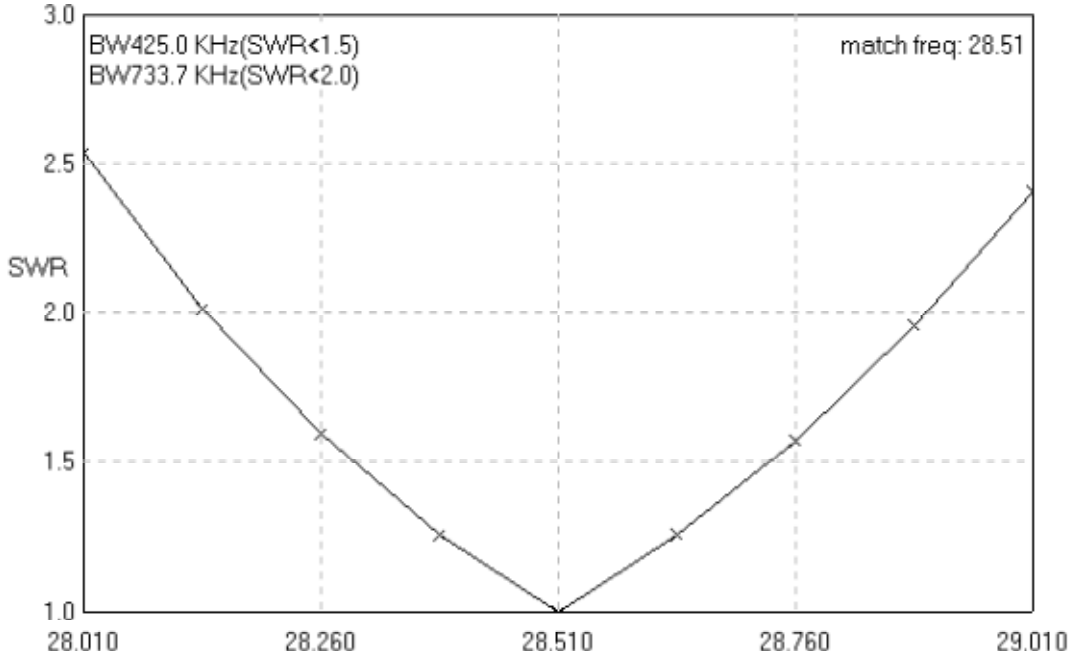


Figure 19 SWR at 50- Ohm coaxial connected through matching device to 'wide' antenna

J- Antenna for 160, 15 and 10(FM) meters
Unusual Look to Usual Things

by Valentin Gvozdev , RU3AEP, http://www.vgvozdev.narod.ru/
gvozdev@excite.com

Introduction

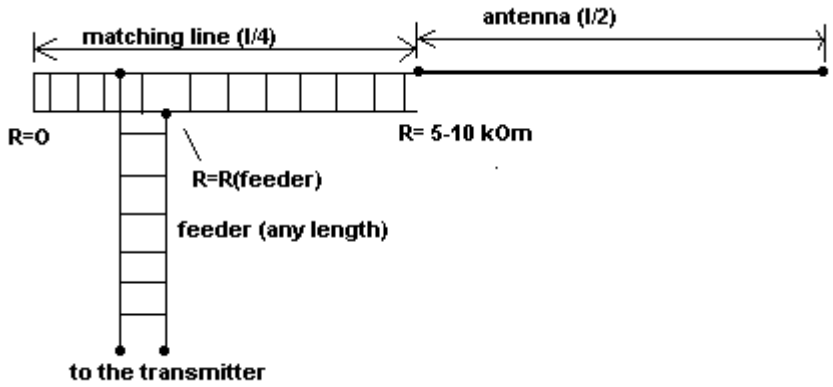
After getting my first amateur license I had to think, what antenna to build for a top-band (160 m), I realized, that conditions are too bad for it. I live in a 7-floor house, which has a roof with a high slope (about 35-40 degrees), which is very dangerous to operate on it. Also, the house is almost completely surrounded by wide streets and electrical wires going along them. After long thinking, I concluded, that there is only one possibility to make an antenna - to hang up a long wire from my roof to the roof of another house. Unfortunately, any dipole-type antenna was unacceptable, because in this case my apartment would have been too far away from the feed point of the antenna, and the condition of right angle (90°) between feeder and antenna itself could not be satisfied. Fortunately, in that time I have read about one very old, but not frequently used antenna - so called Zeppelin-antenna with a matched feeding.

Classical design with an opened line

Actually, this is shortly described in well-known book ("Antennenbuch"), written by DM2ABK (Karl Rothammel), but has been recently developed by Sergey Makarkin (RX3AKT), a radioamateur from Moscow, who has published a good article in "Radio-Design" journal (N2, 1998).

Classical design is presented below (**Figure 1**). As it can be seen, there is feeder with rather high impedance (~300-600 Ohm), and 1/4-wavelength matching line. From one end, this line is shortened, and here its impedance is just a zero (current is high, but voltage is almost zero). Another end of this line is connected to the long wire, which has length exactly 1/2 wavelength. At this point, the impedance is very high (several kiloohms). That is why, a big voltage exists here during a transmission. This is quite suitable for a wire feeding, because a 1/2-wavelength has high impedance when fed from the end.

Figure 1. Classical Zeppelin-antenna design



The feeder from the transmitter with a specific impedance R_f is connected to the matching line in the point, where impedance of the latter is equal to that of the feeder. Such point is usually located not so far from the shortened end. If everything is done properly, feeder may have any length and SWR is closed to 1:1 in rather narrow band, central frequency of which is determined by the geometrical size of matching line and antenna.

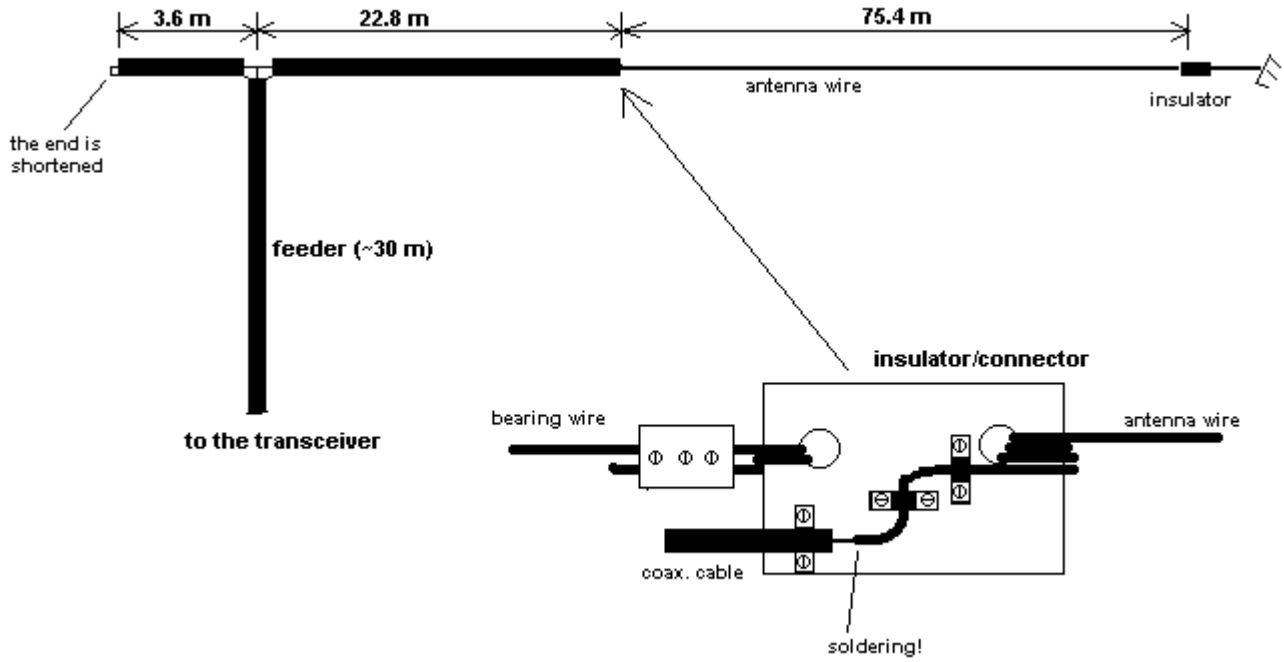
Classical design with a coaxial cable for 160 meters

This design can be used almost without change, but instead of symmetrical feeder a coaxial cable can be used to connect the whole system to the unsymmetrical output of the transmitter (**Figure 2**). Using of a coaxial cable instead of an open line has one big advantage – in contrast with the symmetrical transmission line it is almost insensitive to the environment, weather conditions and can be placed really everywhere.

Such antenna with feeding 'from the end' is much more easy to make, that a simple dipole. Here, antenna wire bears only itself, and this reduces the mechanical strength and thickness of the wire to be used. Also, you may use your window as one the point of antenna fixing. In this case, all the cable will be inside your shack and antenna could be tuned precisely in comfortable conditions. If the beginning of antenna is outside the apartment, most part of matching line can be used as the continuation of the feeding cable.

On **Figure 2** there is a design, that I implemented for using on 160 m amateur band, and which, to my mind, is a perfect solution for the people, who cannot mount a classical dipole.

Figure 2. Long wire antenna for 160 m with a coaxial matching line.



In my case, all coaxial cables have 75 Ohm impedance, the antenna wire, as well as two bearing wires are made from very hard bimetallic insulated cable (outer diameter is about 3 mm). The trickiest part - the connector between cable and antenna - is shown on **Figure 2.**. It should be noted, that voltage on it is quite high, and so everything should be well insulated from each other. It is good idea to place this connector somewhere indoors, otherwise rains and snow may cause decreasing of insulation efficiency and antenna performance. This antenna uses a tuned line made from the coaxial cable, and for proper operation of the whole system the antenna wire should have the length equal to the $\lambda*0.95/2$, and the coaxial line must resonate on the working frequency.

It is a good idea, to connect the shortened end of the matching line to the ground (cold water pipe, heating system, building elements etc.) to provide adequate safety and to reduce possible TV/RF interferences while transmitting.

Tuning and adjusting of the antenna

To achieve what was declared in the previous paragraph, first of all the precise length of the matching line should be determined. Theoretically, it should be closed to $\lambda/(4*\mathrm{sqrt}(d))$ (sqrt - Square Root, d - dielectric constant of the insulator used in the coaxial cable). SQRT(d) value is typically about 1.52 for most cables with polyethylene-based dielectric, that is why, 'shortening coefficient' is about 0.66. But the practical value will be a little different from that.

The lengths indicated on **Figure 2** are mine values, and they can be used as the approximate reference. Exact numbers depends on the antenna environment and should be determined experimentally. It should be noted, that in 'ideal' case it is not a simple task, because in such system three values have to be varied (one is antenna length, and another two are lengths of the parts of the matching line). But as it appeared from my experience, for practical purposes the most important thing is to choose correct total length of the matching line, which must resonate on the desired frequency.

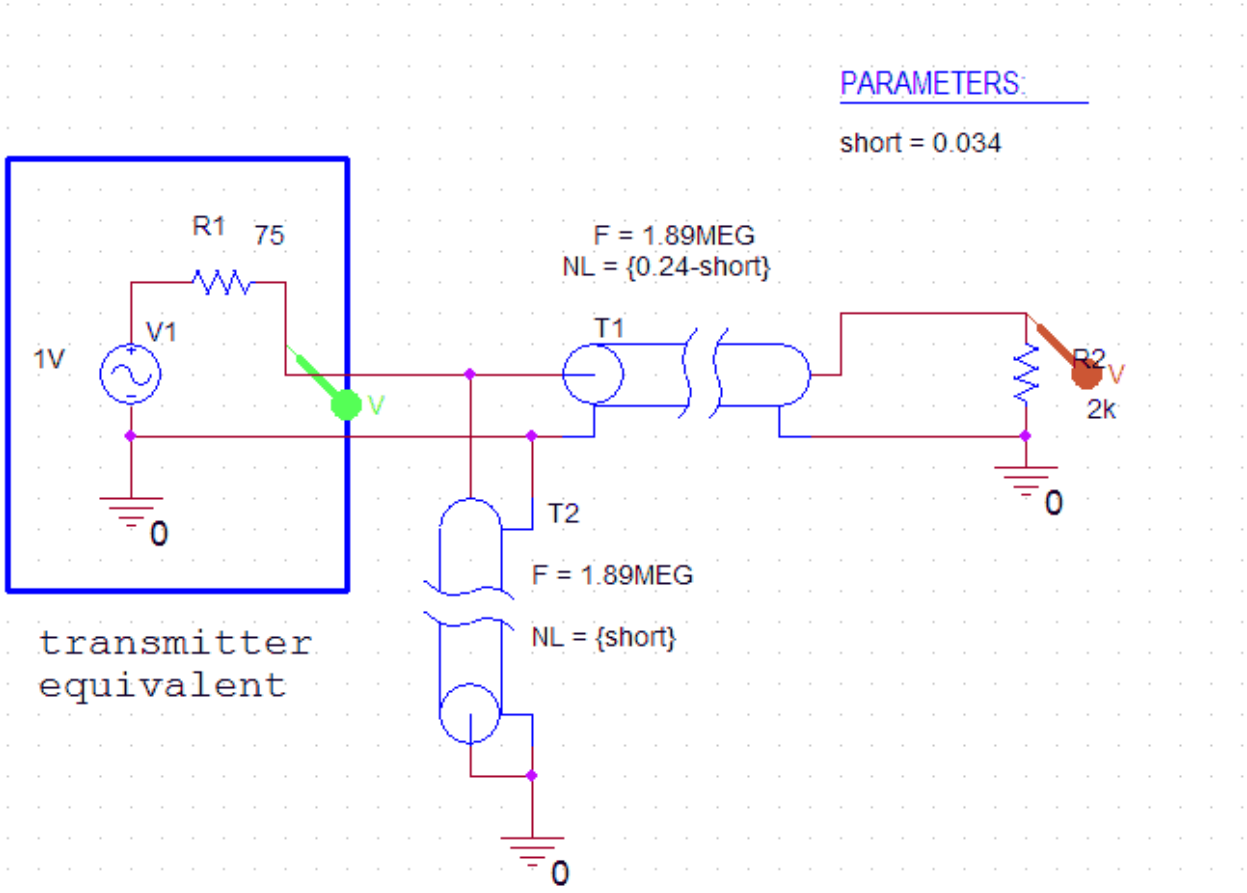
To do this, I suggest to use the following technique. To make your line resonate on the middle of the band (1890 kHz), you first have to make the line about 1 m longer, that estimated length of the tuned line (for example, 24 m), making shortened segment about 3.6 m. Then, connect the 2-3 kiloohms resistor to the "open' end of the line, and

the transceiver through SWR meter - to the feeder. The resistor here serves as a loading instead of the antenna wire.

After assembling of the system, put RF power (1-2 W or even less is enough) on some frequency inside 160 m band into line and watch the SWR. If the line is completely out of resonance, SWR will be closed to infinity, and no power will be dissipated on the resistor. Then, the frequency should be found, which gives the sharp minimum of the SWR. It has to be around 1800 KHz. Here, the SWR is usually less than 1.5:1, and the full power of the transceiver is dissipated on the resistor, which means, that the matching line works well. When touching the 'hot' end of the loading, it may be seemed, that is really very hot – this is due to the high HF voltage, which causes skin burning (be careful to do it, even by low power of RF source!).

For better understanding of these processes is useful to look to the results of simulation of this system using Pspice simulation software. The equivalent schematics (**Figure 3**) includes voltage source V1 in series with 75 Ohm resistor (which emulates output resistance of the transmitter), two coaxial lines T1 and T2 and loading R2. Since Pspice does not allow to set lengths of the transmission lines directly in length units, they are set in wavelengths (NL) on the specified frequency (in our case, F=1.89 MHz).

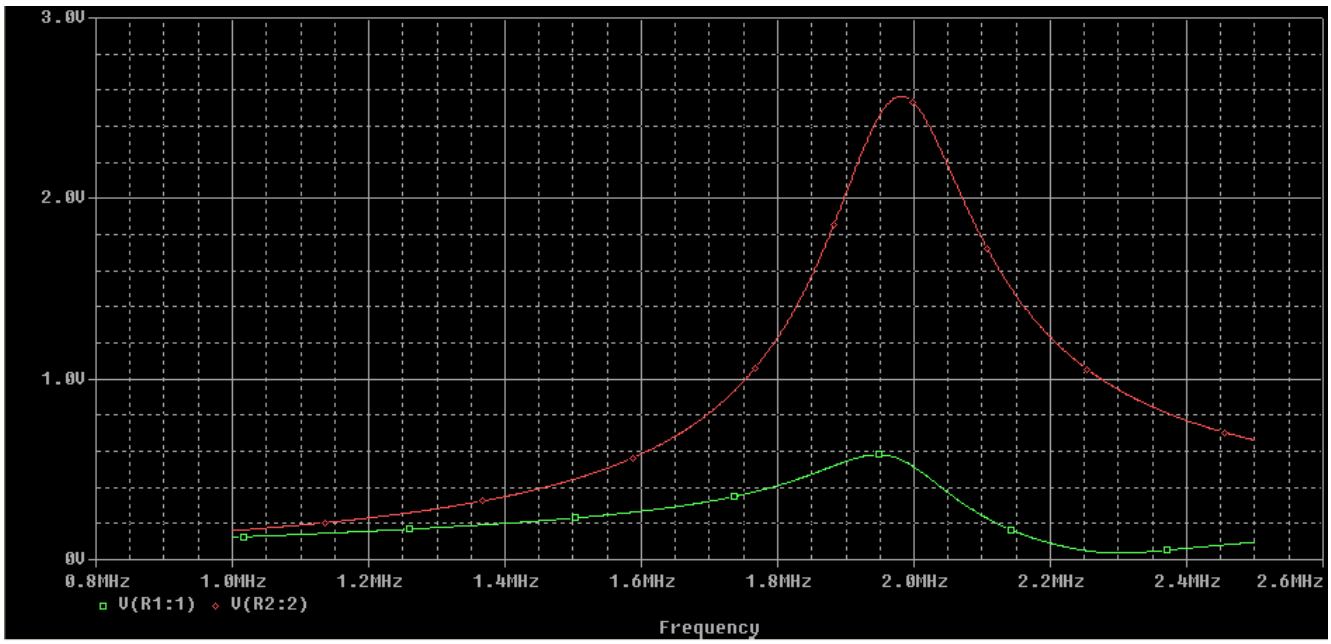
Figure 3. Equivalent schematics for matching line, used for simulation.



Calculated frequency response is presented on the **Figure 4**. Here, the colors of the traces correspond with the colors of the voltage markers on the schematic. As it can be seen, on the resonant frequency about 1.95 MHz there is sharp voltage maximum on R2 (red trace), which reaches 2.6V – it is about 5 times more, than the voltage on the transmitter's output (green trace). Also it should be noted, that on the resonant frequency voltage on R1 (green trace) is closed to one half of source voltage (in our case, 1V). Practically, it means, that there is good matching between transmitter and the "antenna" and most of generated power is dissipated on the loading.

After the resonance has been found, it should be shifted up to the desired frequency. To do this, the end of the cable should be cut carefully in several steps, watching the resonance frequency each time, which must increase with each cut. After you achieve the desired frequency, the matching line is almost ready, and you can mount the whole antenna system in the chosen place. It should be noted, that the minimum of the SWR in mounted antenna is usually 20-30 kHz down, compared to the value achieved by the tuning on the resistor.

Figure 4. Frequency response of the matching line in range 1 – 2.5 MHz



In my case, the antenna for 160 m band had a minimum of SWR at 1875 kHz (about 1.3:1), on the edges of the band SWR increased to 2.0…2.5:1, since the design is a narrow-band one. Compared to my previous dipole, which hanged on the low height (about 5 meters over the ground) along the building, this antenna exhibited much better transmission efficiency and higher signal to noise ratio while receiving.

The same design for 10 meters - cheap and simple.

About 2 years after getting my first amateur license I upgraded it to the higher license class, which allowed me to operate on 10 meters SSB. In that year, there was a perfect propagation on 10 meters band during the daylight time, and I needed an efficient antenna to work on it. Probably, in some time I will have something like rotable multielement Yagi on my roof, but now it seems to me inaccessible as the Moon due to many factors. After some time I decided to repeat what I built for 160 meters for 10 meters, proportionally reducing all geometrical sizes of the antenna wire and matching line.

Since the wavelength on 28500 KHz is just 10.52 m, a half-wavelength dipole should be about 5 meters, and the total length of the coaxial matching line will be 10.52/(4*1.52) = 1.73 m. The feeder is connected to the line 23 cm away from the shortened end. These sizes are relative small and the whole antenna system may be placed without being mounted on the roof, for example just from your window to the neighboring tree.

I made the antenna from a 2 mm copper wire with a plastic insulators at the ends, using 75 Ohm coaxial cable for feeder and matching line. There was nothing difficult to tune the system - I hanged the antenna across my apartment and adjusted the length of the matching line as described above for 160m design using 1.80 m as the starting value. The only thing that should be noted is that the actual resonance of the line is very sensitive to the length variations, so on the final steps the cable should be cut in 1 cm (!) portions or even less to not miss the desired resonance position. After I hanged the antenna on the designated position, SWR was less then 1.5 on all frequencies ranging from 28200 to 29000 KHz.

This antenna is really very simple and cheap, but nevertheless, I allowed me to establish many connections with Europe and even Far East using just about 10 Watts of power. I really enjoyed working on 10 meters ether in local communications and transnational QSOs, and this was made possible just by several hours of time, dedicated to the antenna building and tuning.

About working on other bands – some facts and theory.

Though LW antennas with a feeding through coaxial transformer, which were described above, seem to be monoband, this appeared not completely true. As I found out, the whole system has many resonant frequencies, and some of them, are inside or near amateur bands and can be used for working on these bands.

As it could be expected, operation on the frequencies, which are twice more that ‘native’ ones, is impossible. When using an antenna for 160 m, on 80 m band observed SWR is closed to infinity and the transmission efficiency is not more that by using a

random wire with length of several meters... Simply it can be understood, that on doubling the working frequency the matching line is completely out of resonance, and works as a “short” for the transmitter. But everything has advantages, and this fact means not only impossibility to work on 80 m, which is definitely bad, but also deep suppression of 2- nd harmonic by working on 160 m, which is really well.

Almost the same situation is on 40 m band. Here the active component of input impedance of the antenna (measured by noise bridge) is also quite low (several Ohms), and no resonance exists inside or near amateur frequencies.

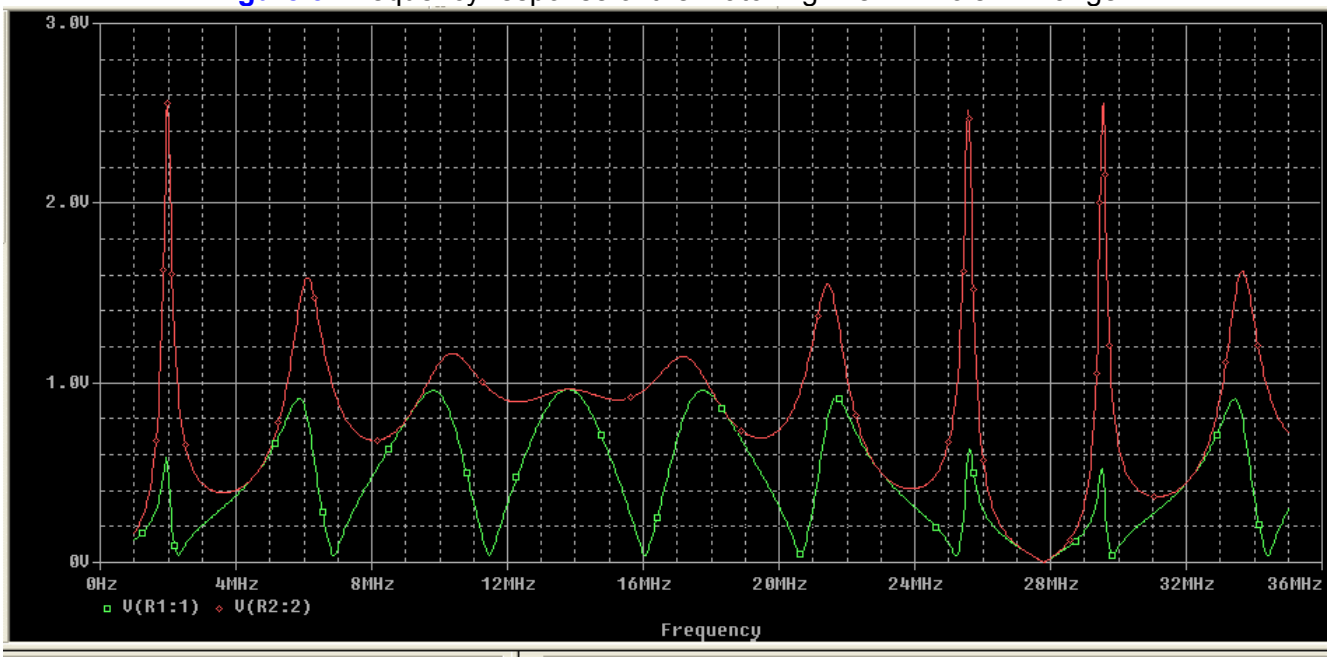
But if you try to work on this antenna on 15 and 10 meters bands, the situation is more optimistic. In my case, on 21430 KHz the SWR was about 1.3:1 and increases to 2.5:1 when moving down to 21000 KHz. Measured impedance was about 55 Ohm with a low capacitive reactance. From first sight, it is quite strange, but nevertheless, antenna behaved well on this band, and using just 10 W of power, I was able to make long-distance QSO’s even with North America. The most interesting fact was, that this was “true”

resonance of the antenna, without any participation of the feeder (SWR did not change significally when the feeder length was alternated).

In contrast to this, on 10 meters band the antenna behaves very poor – the air seems to be “empty”, and even common industrial noise is received with a level comparable to internal noise of the receiver. Compared to the special 10 meters antenna (see above), the signal of distant correspondents were weaker by 10-20 dB (!), and on transmission even my neighbors gave me reports like 53-54. However, when frequency was moved up to 29 MHz and higher, the efficiency improved dramatically.

To understand this phenomenon, some calculations were performed. First of all, it was found, that frequency response of the matching line with a resistive loading (see **Figure 3**) in range 1.5 – 32 MHz has many maxima, and one of them is inside 15-m amateur band (**Figure 5**, red trace). Another maxima is near 29.5 MHz – in the upper part of 10-m band.

Figure 5. Frequency response of the matching line in whole HF range



I guess, that these results may be assumed at least as a qualitative explanation of the antenna behavior. I say “qualitative” because the whole system can not be adequately represented by a matching line with a resistor at the end – impedance of the antenna wire also should be taken into account. However it is clear, why besides ‘native’ band, antenna works well on 21 MHz, and why on frequencies about 28500 there is a minimum of performance, which rapidly increases when moving up to 29 MHz.

Conclusion

As a conclusion is can be said, that LW antenna with a coaxial matching line (J-antenna), which is designed for 160 m band, can do perfect job on 15 meters and on a part of 10 meters band also without any switching and tuning devices. Of course, the efficiency on ‘upper’ bands is be substantially lower, that on ‘native’ one due to RF losses in the matching line (which actually works with a very high SWR). But to my mind it is still acceptable, especially in the case, when there are no conditions to mount huge and efficient antennas.

MULTIRANGE TRAP ANTENNAS

Multirange trap antenna: history and fundamentals

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Recently multirange trap antennas are widespread among radioamateurs. As matter of fact, the type of antennas was invented in the USA by H. K. Morgan, US patent # 2229856, 1938 (by reference **[1]**). Probably the first article about a trap antenna was published in reference **[2]** at 1940. So, what is the antenna and how is it work? Let's see it on the example of a ham vertical trap antenna in order to simplify a problem. **Figure 1** shows us a schematic of such antenna.

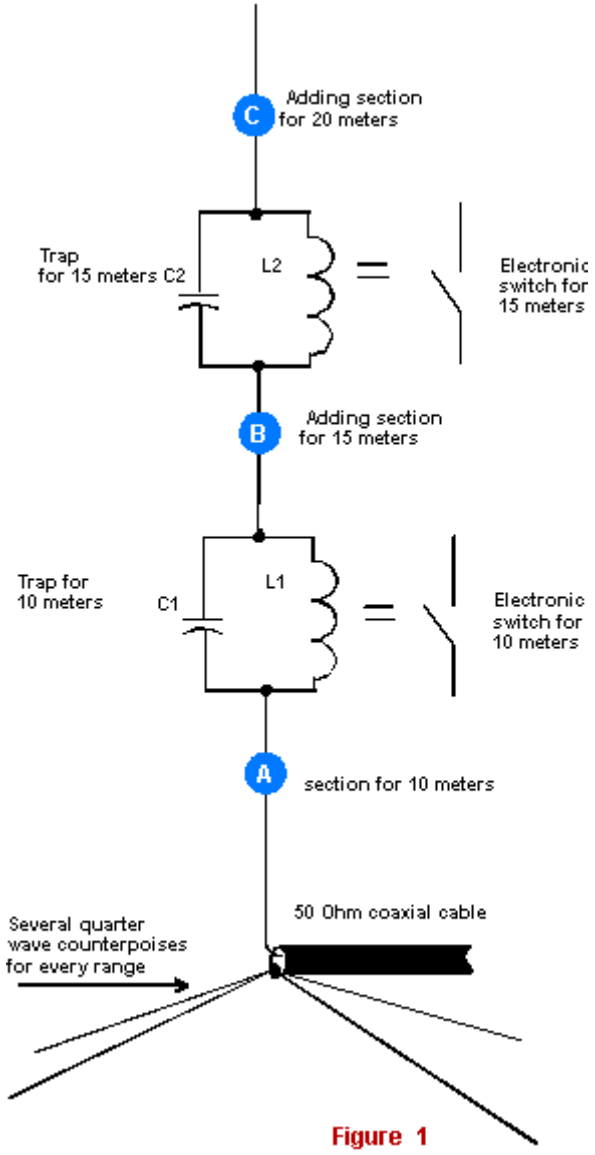


Figure 1 A ham vertical trap antenna

10 meters: Section A is tuned for operation on 10-meters by its length. Trap L1C1 turn off upper antenna parts behind the trap from operation of the antenna when 10 meters range is used.

15 meters: By length of the Section B we tune the antenna parts "Section 1 plus L1C1 plus Section B" to resonance to 15-meters. Trap L2C2 turn off upper antenna parts behind the trap from operation of the antenna when 15 meters range is used.

20 meters: By length of the Section C we tune the antenna parts "Section 1 plus L1C1 plus Section B plus Section C" to resonance to 20-meters.

And so on for other ranges: In the similar way the antenna would be tuned for others ham HF- ranges. You see, it is possible to do an antenna for any number of HF- ranges! But there are several lacks. Upper parts of the antenna behind a proper trap do not use (or, practically do not use) for radiation. Another lack is that the antenna wire is broken at several places by trap circuits. Every trap circuits should be tune in to own resonance frequency. Trap circuits must have high temperature stability, because the antenna is used at the open air. Traps work at a resonance mode so a high level of RF voltage is across trap capacitors at transmission mode. Thereof it needs to use a high quality capacitor for every of the traps.

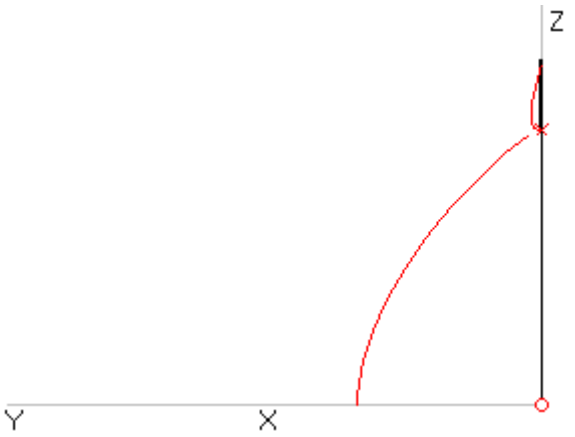
Vertical trap antenna WA1LNQ: One of the most popular sample vertical trap antenna is the antenna WA1LNQ **[2]**. The antenna is used on 10 and 15 meters. **Figure 2** shows the scheme for the antenna.

The antenna made from two insulated from each other metal tubes by length of 240,7 (section A) and 62,9 (section B) centimeters and in OD 18 to 25 millimeters. The length of an insulating insertion is 5,8 centimeters. Over the insulating part is spooled the trap spool. A copper tube in diameter of 3 to 5 mm is used for the spool, and the spool contains 2 turns with step 1 turn on 25-mm of winding. Average diameter of the trap spool is 55-mm. As a trap capacitor is used a length of a 50-Ohm coaxial cable with an initial length equal to 80 centimeters.

Tuning of the Antenna WA1LNQ: At first, tune the antenna in 10-m range. At the tuning the length of the coaxial cable, that makes the trap capacitor, is gradually shortened to minimum SWR in 10 meters. After this, tune the antenna to minimum SWR at 15 meters. It is possible to do by a small changing of the length of the upper section B.

Below you can see input impedance, SWR and DD of the antenna W1LNQ. The figures are obtained with the help of Free Antenna Simulation Program

Multirange Trap Antennas



Current Distribution at 10 meters

Current Distribution at 15 meters

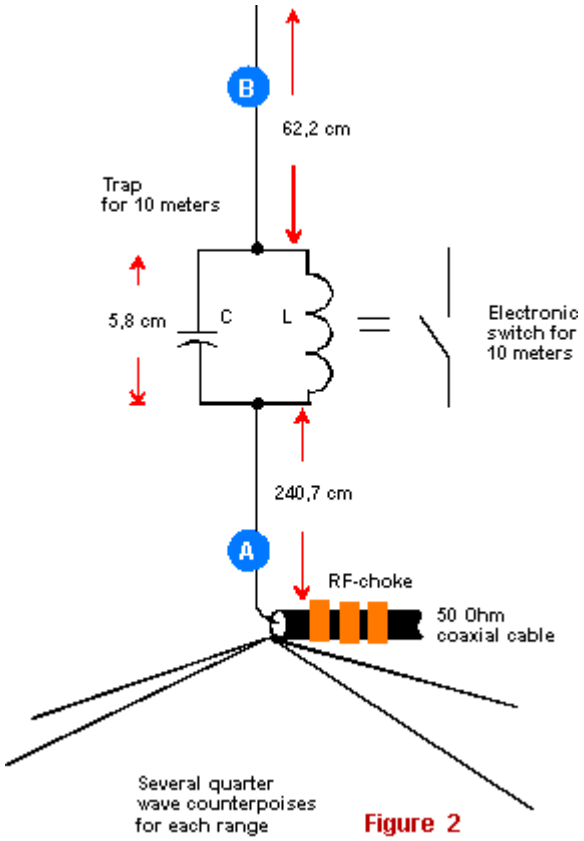
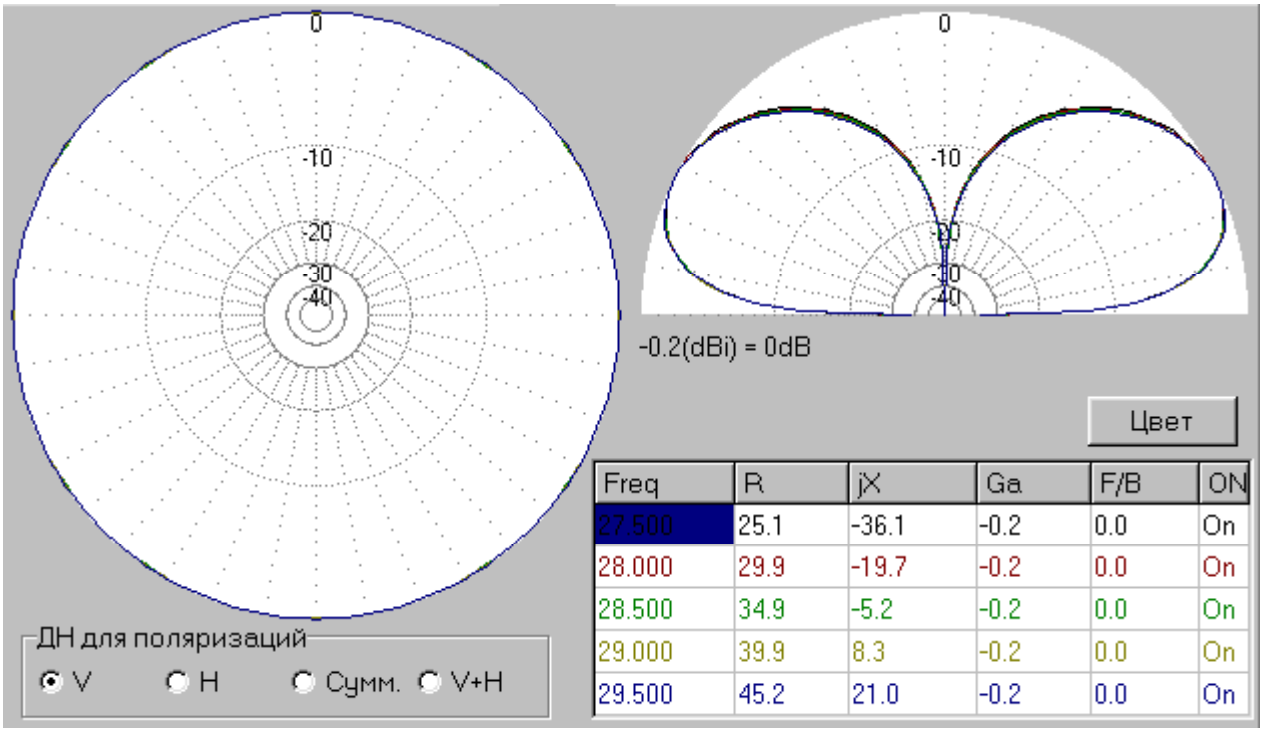


Figure 2 Antenna WA1LNQ

Diagram Directivity and Input Impedance at 10 meters W1LNQ Antenna



Freq	R	jX	Ga	F/B	ON
27.500	25.1	-36.1	-0.2	0.0	On
28.000	29.9	-19.7	-0.2	0.0	On
28.500	34.9	-5.2	-0.2	0.0	On
29.000	39.9	8.3	-0.2	0.0	On
29.500	45.2	21.0	-0.2	0.0	On

MMANA (MININEC based). Section A has diameter of 24 millimeters, section B has diameter of 18 millimeters.

Antenna input impedance. Vertical Trap antenna: You can see that only first antenna part, it is section A at **Figure 1**, has length in λ/4. So, the input

SWR at 10 meters W1LNQ Antenna

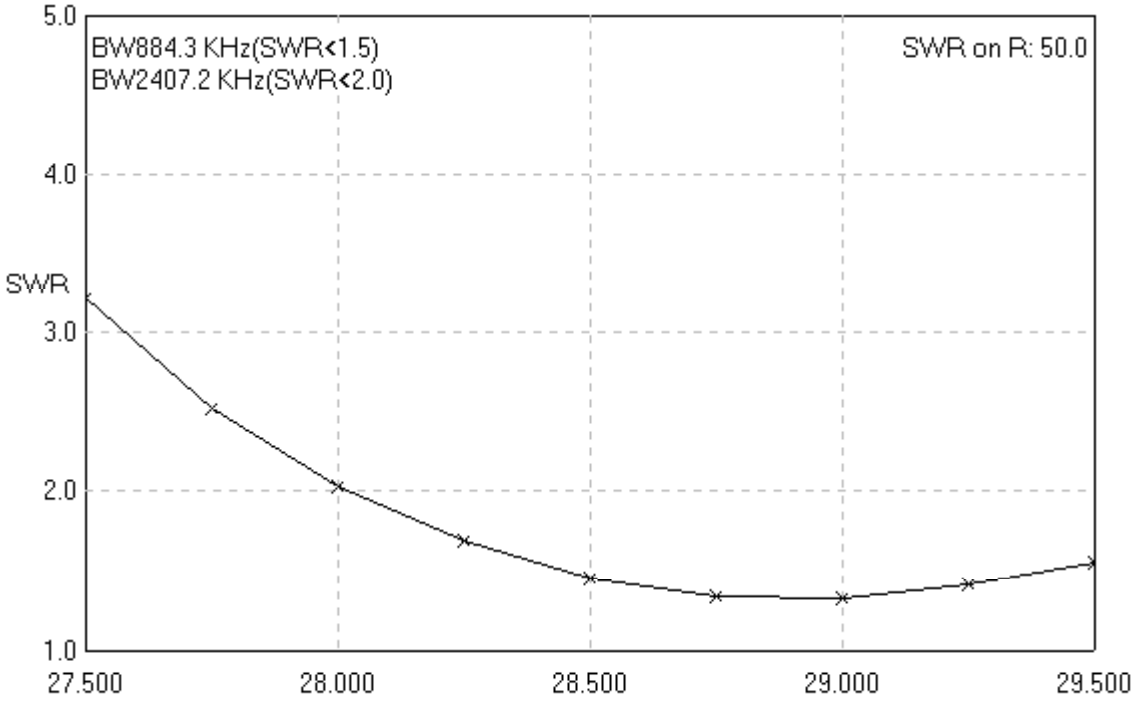
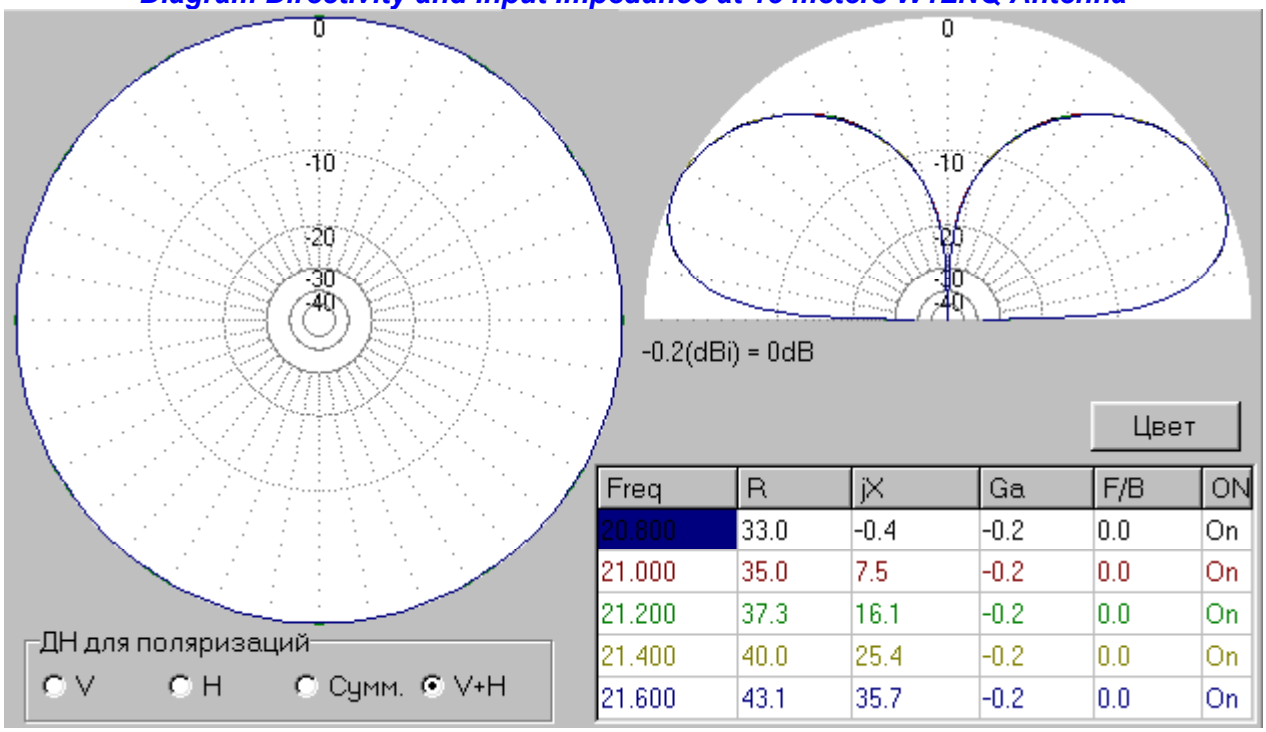


Diagram Directivity and Input Impedance at 15 meters W1LNQ Antenna

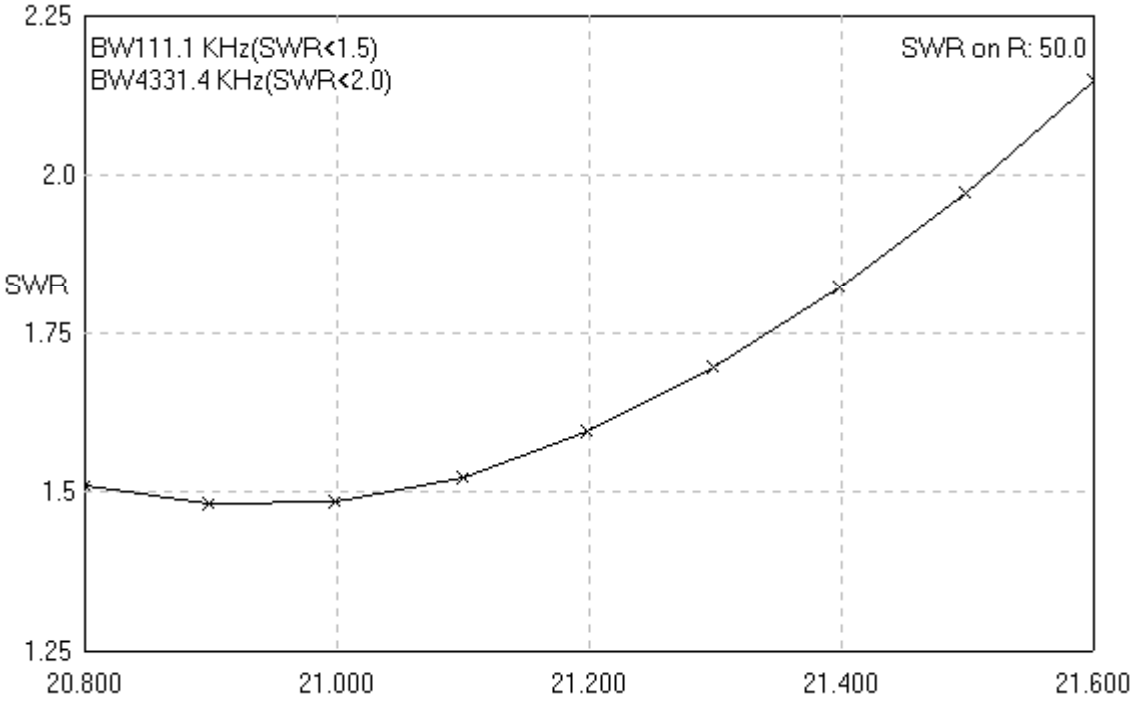


Freq	R	jX	Ga	F/B	ON
20.800	33.0	-0.4	-0.2	0.0	On
21.000	35.0	7.5	-0.2	0.0	On
21.200	37.3	16.1	-0.2	0.0	On
21.400	40.0	25.4	-0.2	0.0	On
21.600	43.1	35.7	-0.2	0.0	On

impedance of the antenna at 10 meters is close to 40-Ohms, and50-Ohms coaxial cable can be used for feeding of the antenna at the range. However, physical length of antenna consisting of another following section plus the previously section (or sections) is less then λ/4. Inductors of the traps work as a lengthening spools for the proper section. Input impedance of the antenna working at lower then 10 meters range is less then 30 Ohms in the theory, but in practice, the input impedance for 15 and 20 meters range is close to 40 Ohms because losses in antenna parts and antenna ground. So, a 50-Ohms coaxial cable can be used for feeding of the antenna at all of the ranges. For a proper work a vertical trap antenna must have several counterpoises for every of operation ranges, especially for low amateur HF ranges 40-, 80- and 160-m.

Antenna input impedance. Dipole Trap antenna: Morgan trap antenna **[1]** was done as a dipole. It is known, that a λ/2 (physical length) dipole antenna has input impedance close to 75 Ohms, see **Figure 3A**. A shortened by a lengthening spool dipole

SWR at 15 meters W1LNQ Antenna



antenna with electrical length in λ/2 (having physical length bit less then λ/2) has input impedance less them 75 Ohms and maybe, close to 60- 50-Ohms, see **Figure 3B**. So, for feeding a dipole trap antenna a 50-Ohms coaxial cable can be used with a high efficiency.

A dipole trap antenna is very easy for tuning and has high efficiency, however, radio amateurs very seldom make as a vertical as dipole trap antenna having a number of traps more than one. The reason is that the antenna sections should be electrically insulated from each other. It is hard enough to do a mechanical strength design of such antenna in radio amateur conditions. Radio amateurs usually prefer a W3DZZ antenna. The antenna has only one trap, and, as it seems by many hams, works at several amateur ranges. What is a W3DZZ?

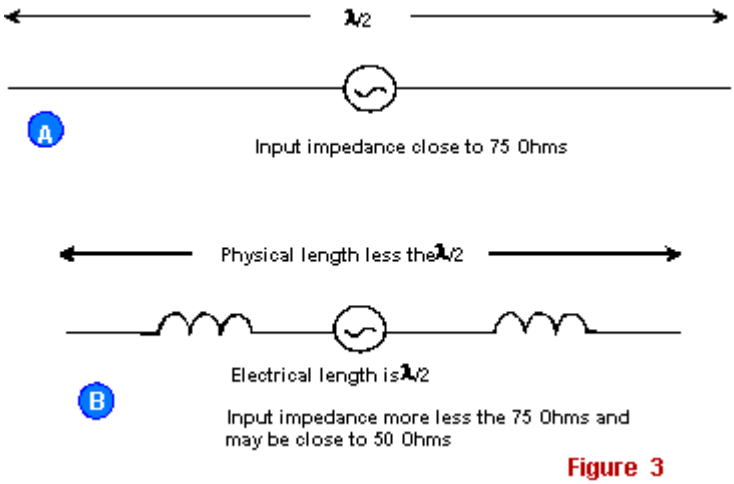


***Figure 3** A shortening and full size dipole antenna*

Antenna W3DZZ: In1955 C. L. Buchanan. W3DZZ, developed a multirange dipole antenna with only one trap, see reference **[4]**. Recently the antenna is known as "antenna W3DZZ." **Figure 4** shows schematic of the antenna W3DZZ.

Antenna W3DZZ works in several amateurs range with low SWR in its feeder. Proper choosing data of the trap turns the trap or to lengthening inductor at low range (ranges) or to shortening capacitor at high range (ranges), or to only a trap at a proper range. For the antenna shown at **Figure 4**, trap LC is the trap for 40

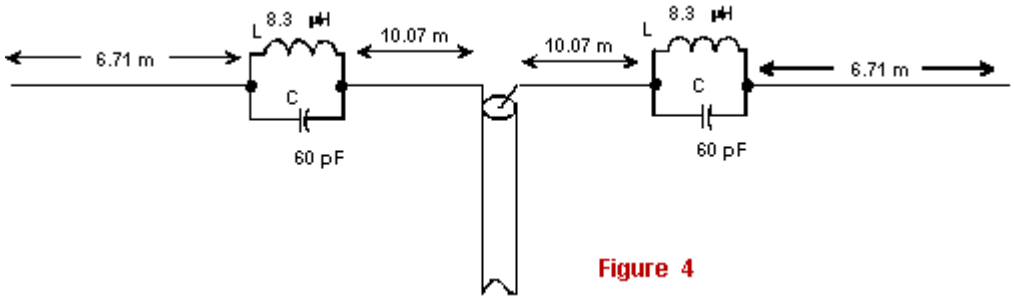
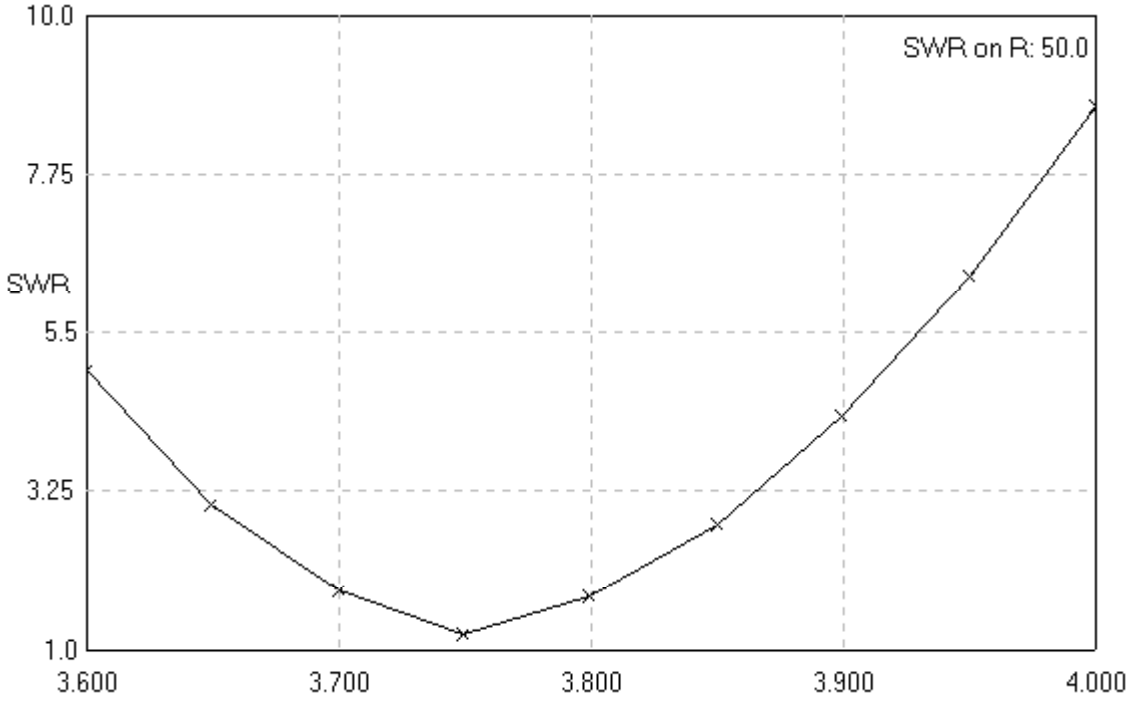
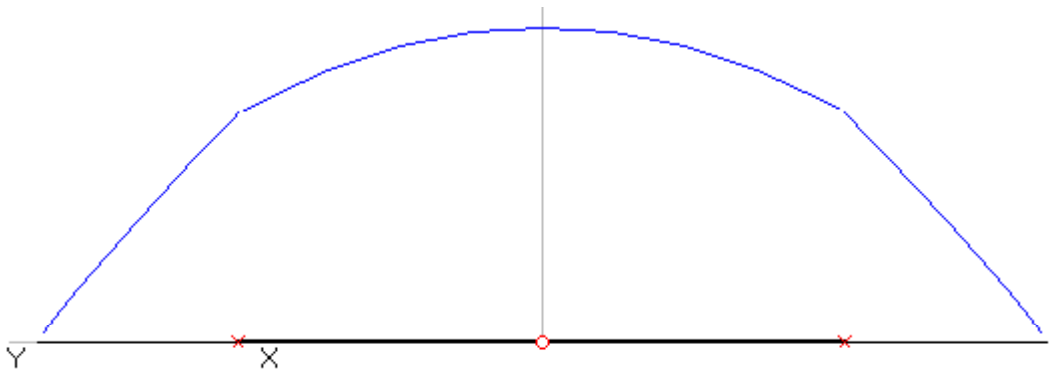


Figure 4 *Antenna W3DZZ*

meters, lengthening inductor for 80 meters, and shortening capacitor for 20, 15 and 10 meters. The antenna (**Figure 4)** does not work at WARC bands. However, the antenna does not work properly at 20, 15 and 10 meters. You can see data obtained with ***Free Antenna Simulation Program MMANA*** ***(MININEC based)*** for the W3DZZ (see **Figure 4**). Antenna wire has diameter of 2 millimeters. You can see, that a SWR at 20, 15 and 10 meters is too high. It is impossible to find such length of the antenna and data for trap that the antenna works at all of the ranges! So, an ATU and a good coax is need for the antenna if you work at 20, 15 and 10 meters.

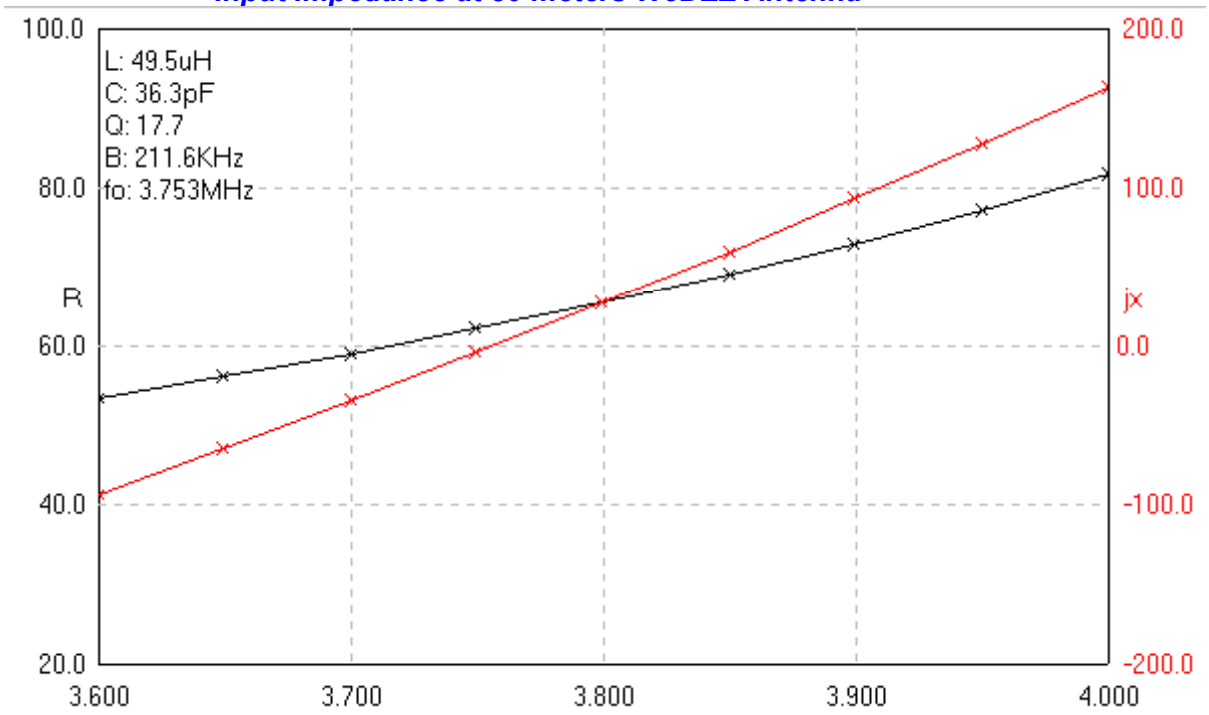
SWR at 80 meters W3DZZ Antenna



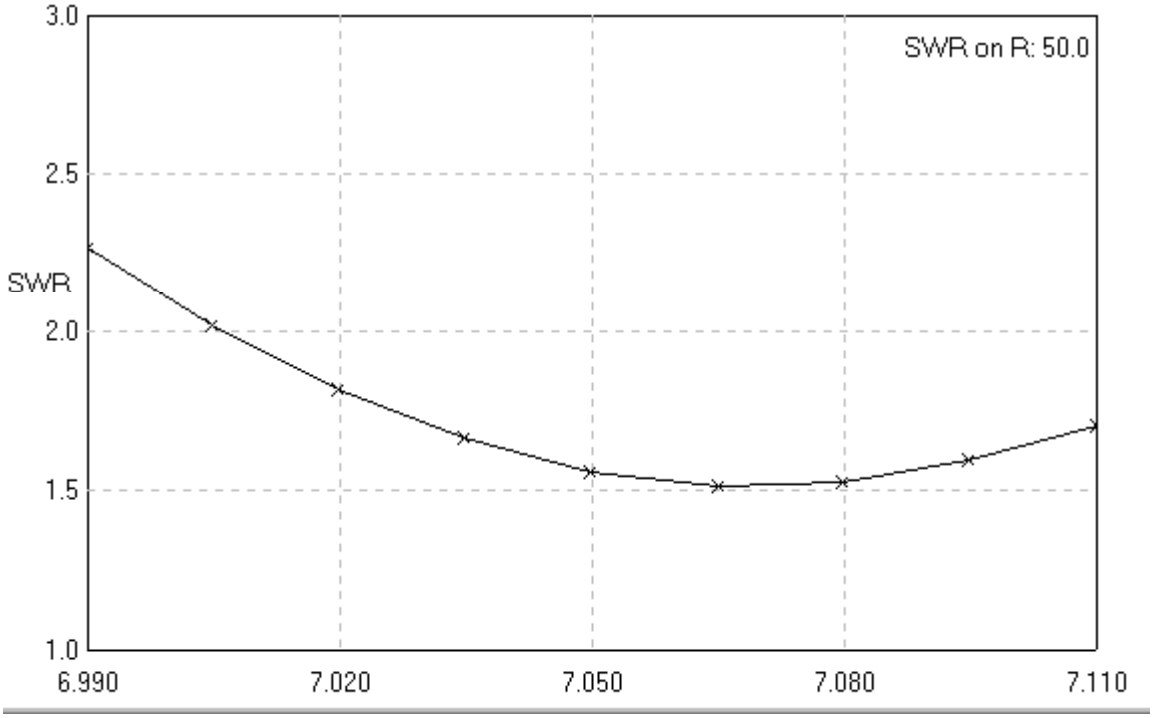


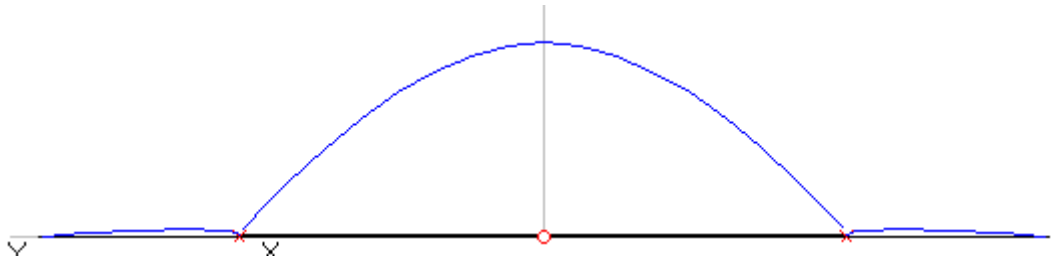
Current Distribution at 80 meters W3DZZ Antenna

Input Impedance at 80 meters W3DZZ Antenna



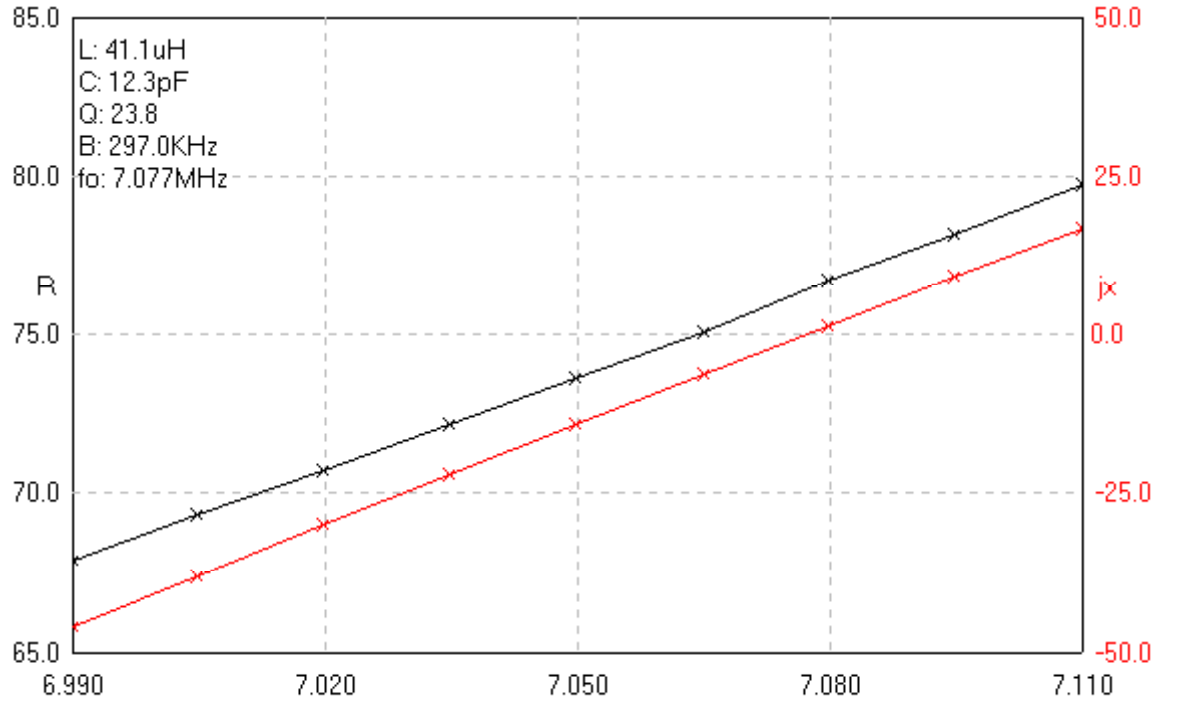
SWR at 40 meters W3DZZ Antenna



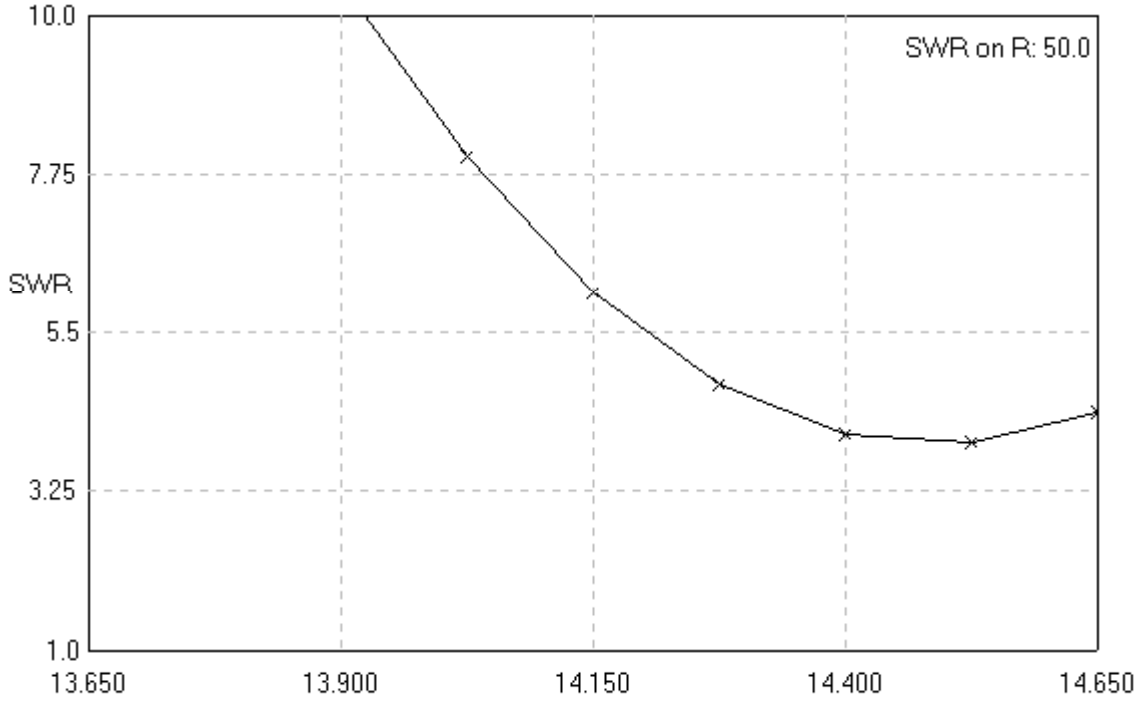


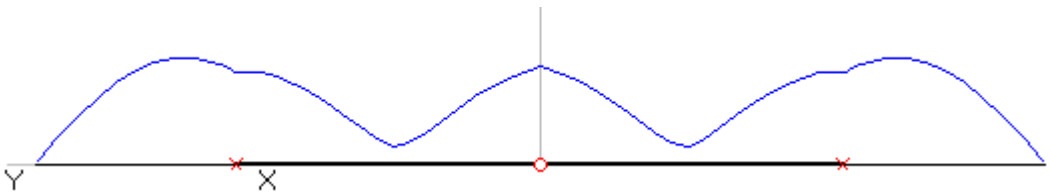
Current Distribution at 40 meters W3DZZ Antenna

Input Impedance at 40 meters W3DZZ Antenna



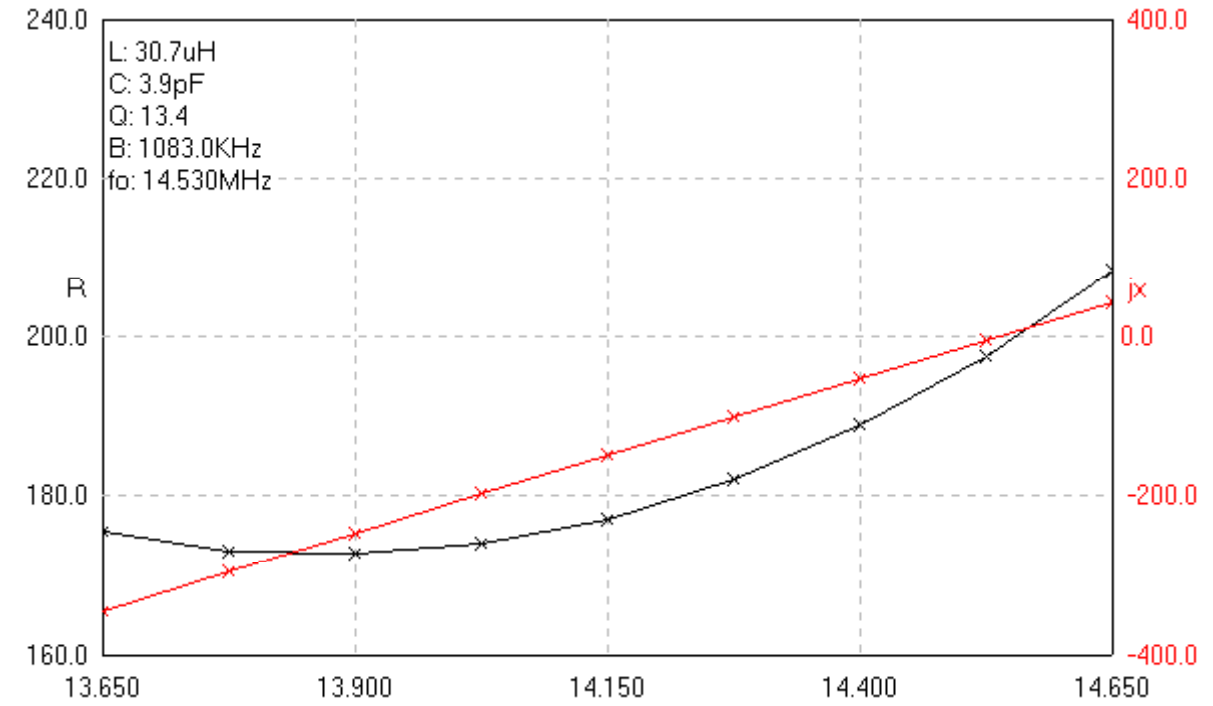
SWR at 20 meters W3DZZ Antenna



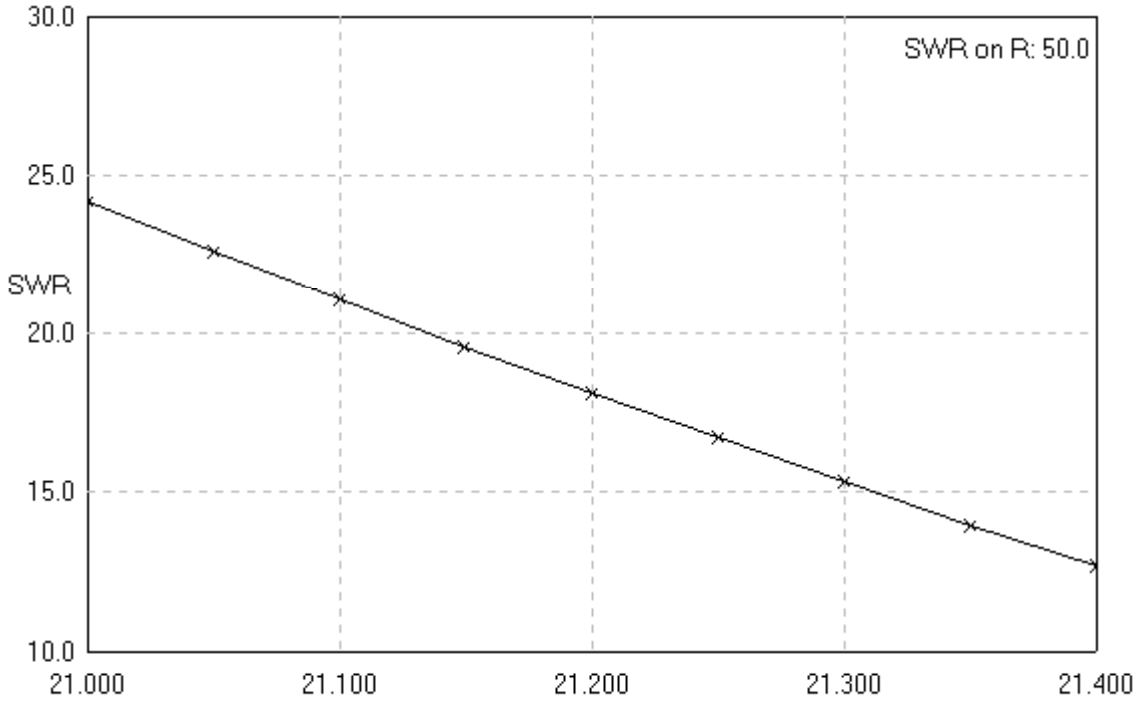


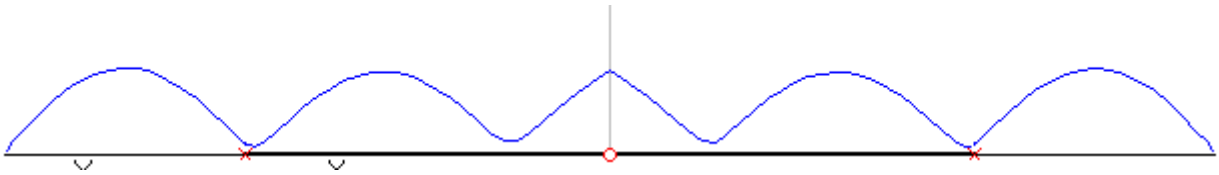
Current Distribution at 20 meters W3DZZ Antenna

Input Impedance at 20 meters W3DZZ Antenna



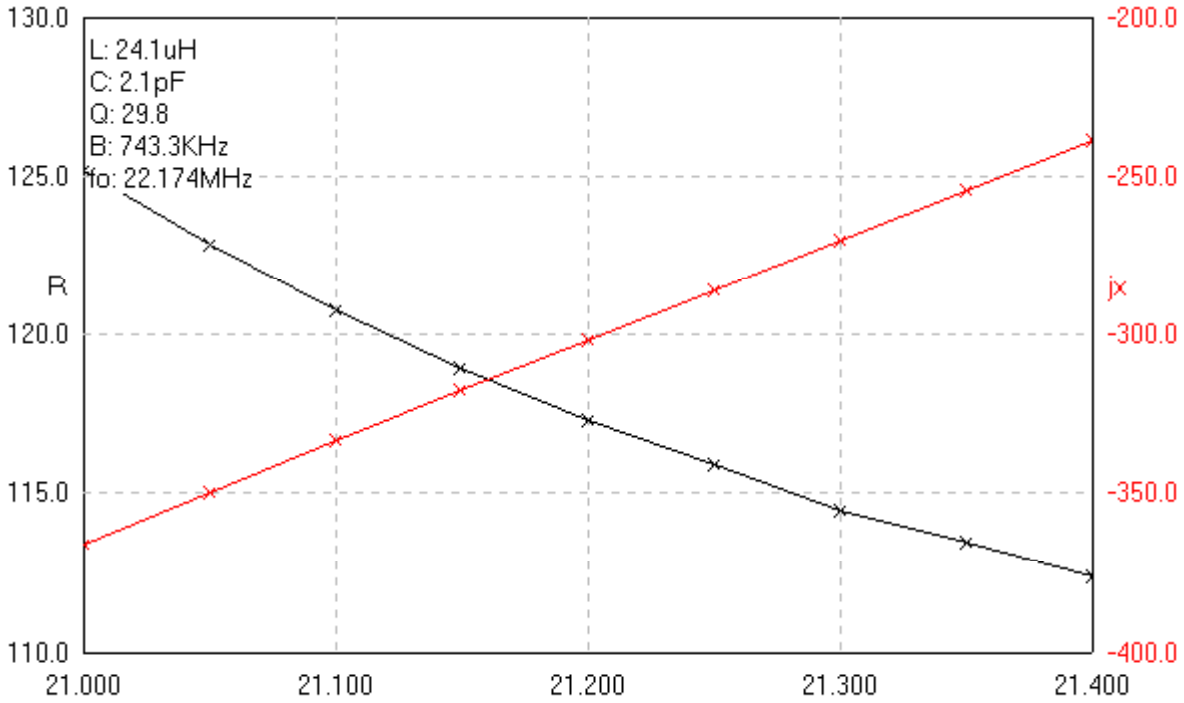
SWR at 15 meters W3DZZ Antenna



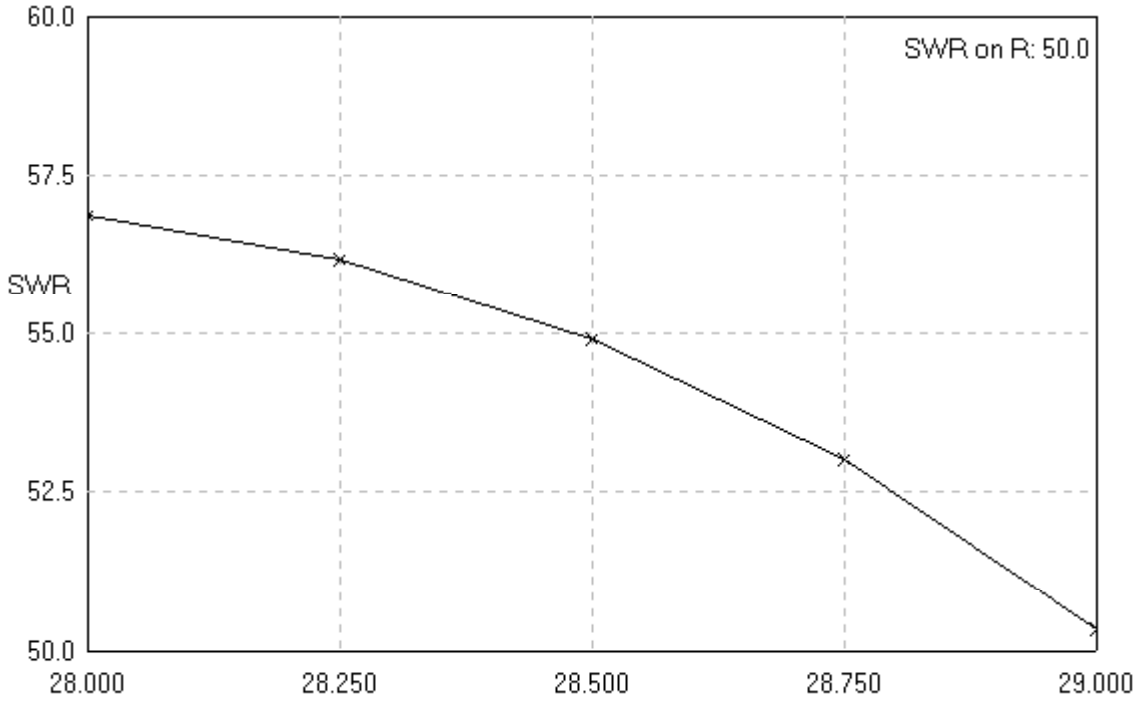


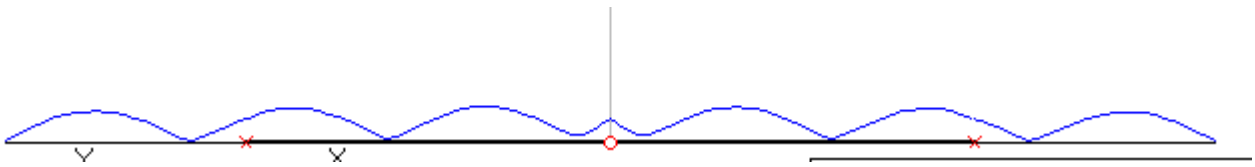
Current Distribution at 15 meters W3DZZ Antenna

Input Impedance at 15 meters W3DZZ Antenna



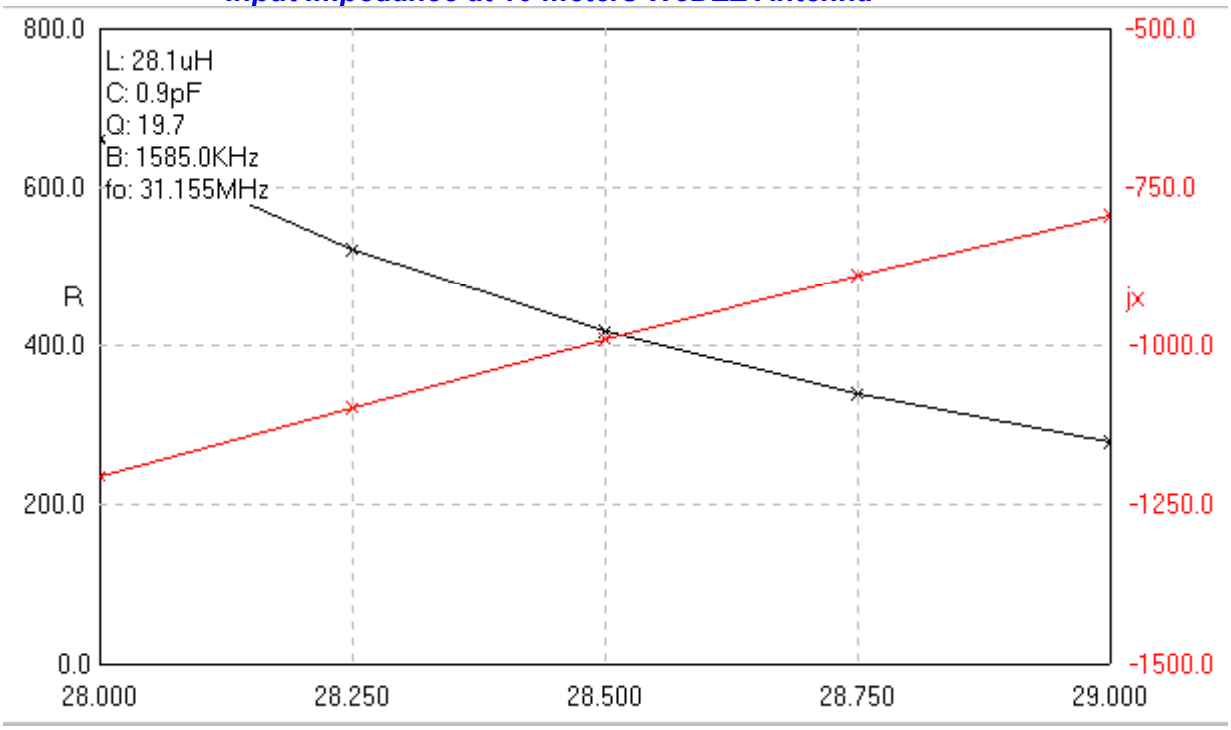
SWR at 15 meters W3DZZ Antenna





Current Distribution at 10 meters W3DZZ Antenna

Input Impedance at 10 meters W3DZZ Antenna



Antenna W3DZZ has input impedance close to 60 Ohms at 80 and 75 at 40, , so, a 75-Ohms coaxial cable can be used for feeding of the antenna.

Hams make antenna W3DZZ also in a vertical installation, where the antenna has input impedance close to 30- 40 Ohms (in twice less the dipole design), so a 50-Ohms coaxial cable can be used for feeding of the antenna at all of the ranges. For a proper work a vertical trap antenna must have several counterpoises for every of operation ranges, especially for low amateur HF ranges 40-, 80- and 160-m.

LC trap design: Trap spool has 8.3-μH and contains 19 turns of silvered copper wire of diameter in 3-mm. Diameter of winding is 50-mm. Length of winding is 80-mm. The trap should be tuned to resonance to the frequency 7,05 (7.2 for USA) MHz. It is possible use a GDO for the tuning. A capacitor at 3-pF is bridged to trap capacitor when the trap is tuning to the resonance. The capacitor is simulated a stray capacitance of the antenna sections.

Antenna tuning: At first, with the help of a GDO tune trap to 7,05 (7.2 for USA) MHz. Trap is tuned separately from antenna. At second, get a minimum SWR on 40 meters by length A. At third, get a minimum

SWR on 80 meters by length B. At thus, you can get SWR (well, see in the above figures, the SWR is not so at 20, 15 and 10 meters.

Four band vertical one-trap antenna:

Hams often use a shortened sample of the W3DZZ antenna intended for 40, 20, 15, and 10 meters. At radio amateurs literature there are several description of the antenna, as at dipole as at vertical installation. However, the first description, which I found off for a vertical four band trap antenna, was made by K2GU in reference **[5]**. **Figure 5** shows the schematic of the antenna.

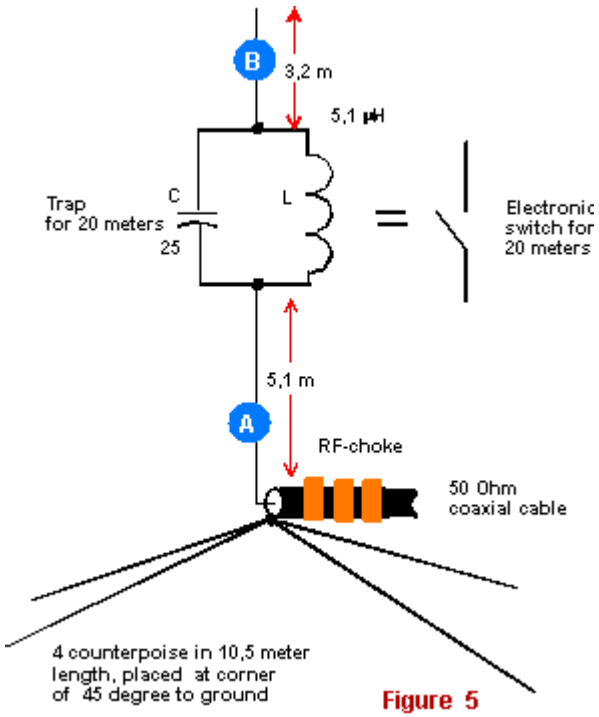


***Figure 5** A four-band trap vertical antenna*

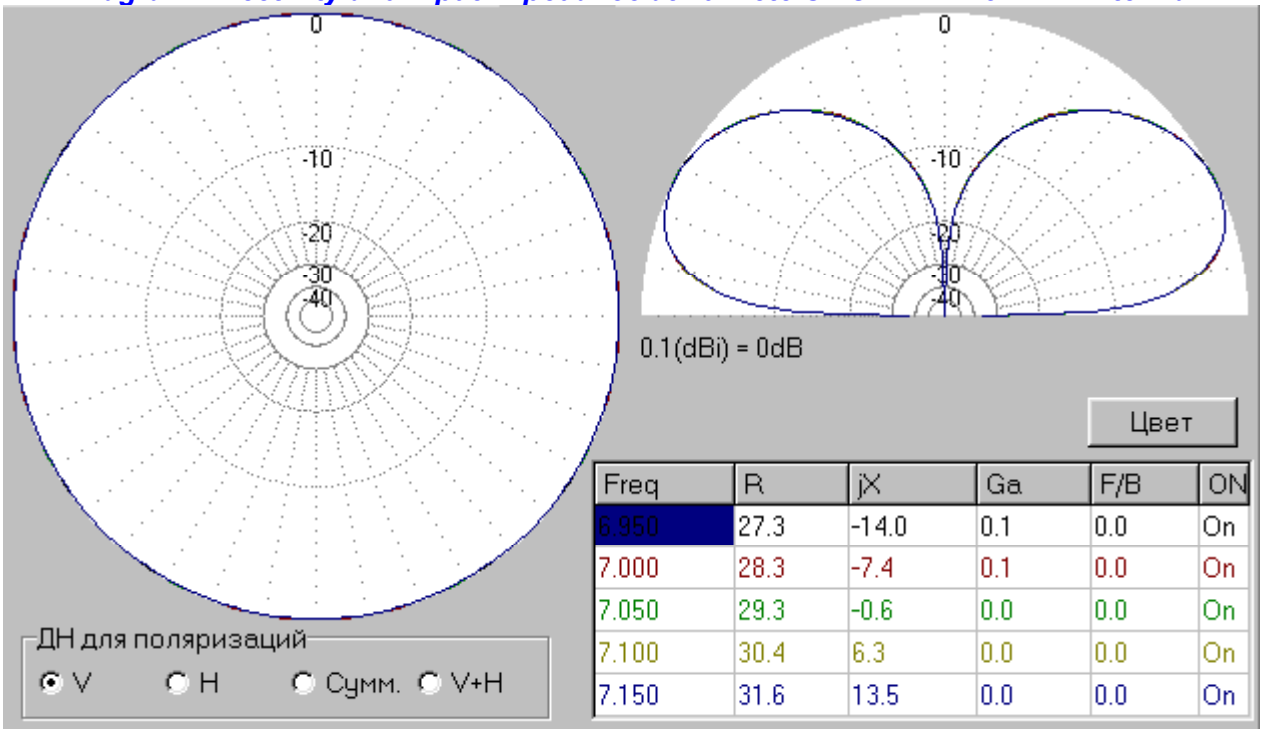
20 meters: Section A is tuned for operation on 20-meters by its length. Trap LC turn off upper antenna parts behind the trap from operation of the antenna when 10 meters range is used.

40 meters: By length of the Section B we tune the antenna parts “Section 1 plus LC plus Section B” to resonance to 40-meters.

15 and 10 meters: The trap serves as a shortening capacitor at that ranges.

Below you can see input impedance, SWR and DD of the antenna W1LNQ. The figures are obtained with the help of Free Antenna Simulation Program MMANA (MININEC based). Section A has diameter of 20 millimeters, section B has diameter of 10 millimeters. You can see, that a SWR at 15 meters is too high. It is impossible to find such length of the antenna and data for trap that the antenna works at all of the ranges. So, an ATU and a good coax is need for the antenna if you work at 15 meters. A 50-Ohm coaxial cable can be used for feeding of the antenna at all of the ranges.

Diagram Directivity and Input Impedance at 40 meters FOR 4B- W3DZZ Antenna



Freq	R	jX	Ga	F/B	ON
6.950	27.3	-14.0	0.1	0.0	On
7.000	28.3	-7.4	0.1	0.0	On
7.050	29.3	-0.6	0.0	0.0	On
7.100	30.4	6.3	0.0	0.0	On
7.150	31.6	13.5	0.0	0.0	On

SWR at 40 meters FOR 4B- W3DZZ Antenna

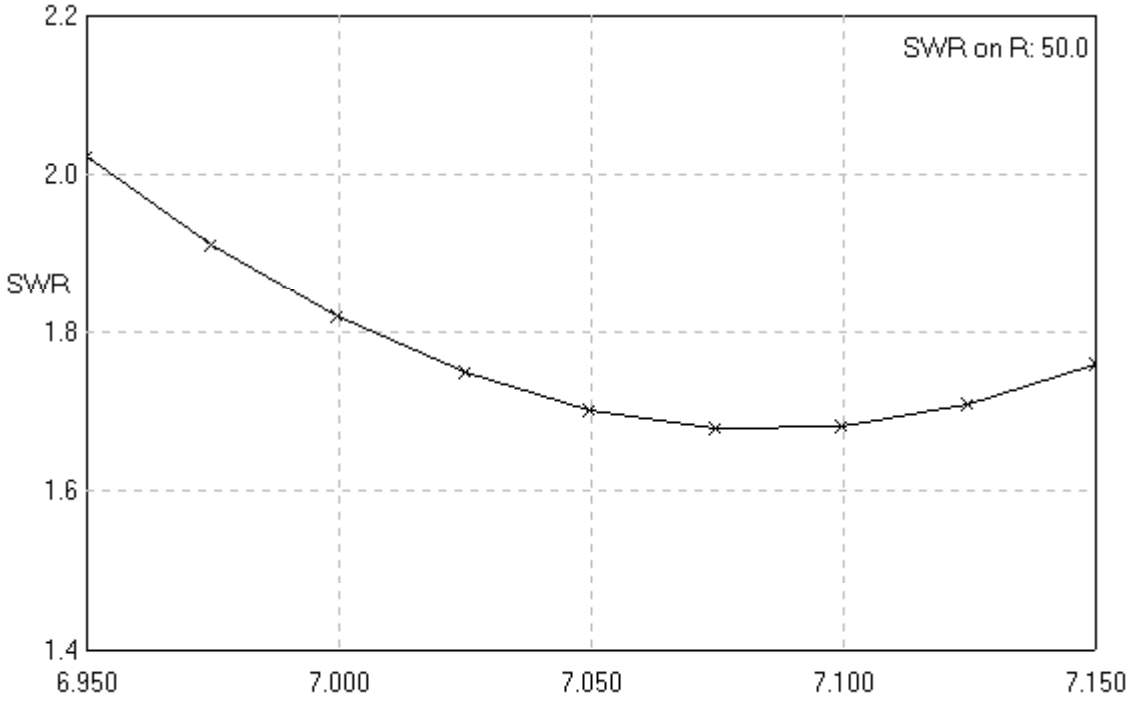
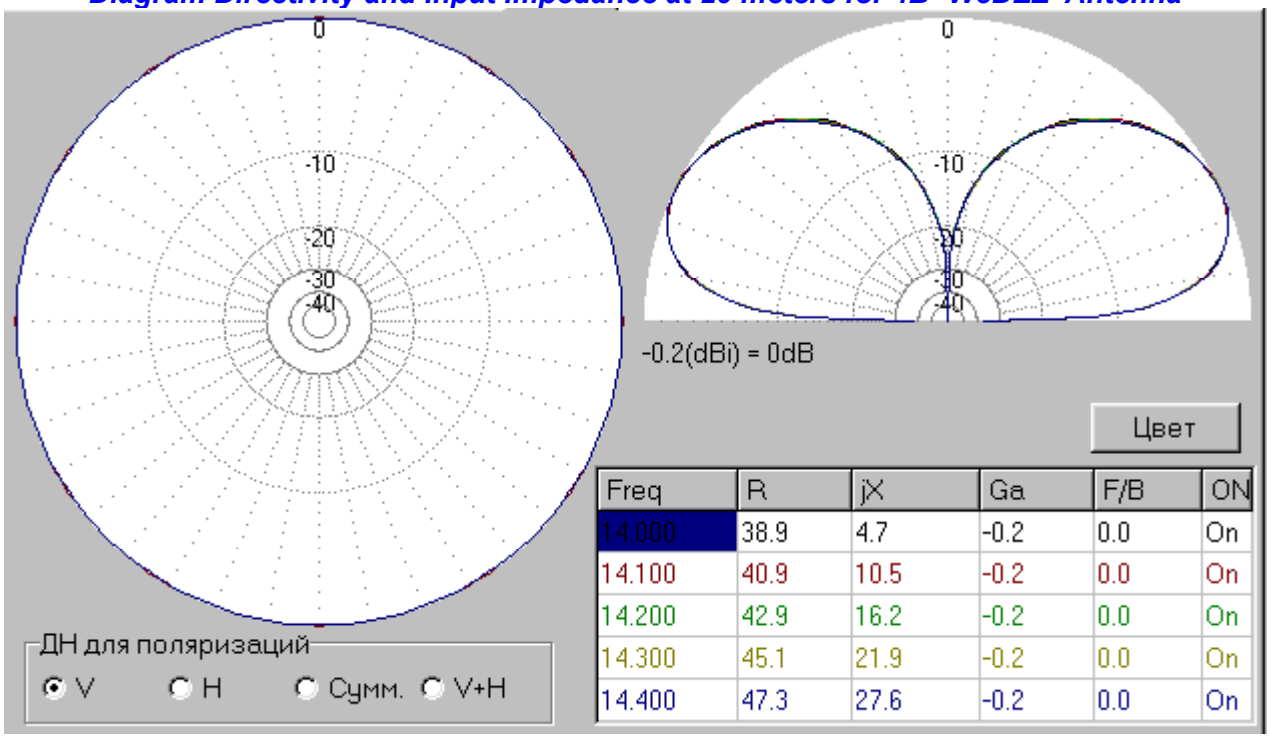
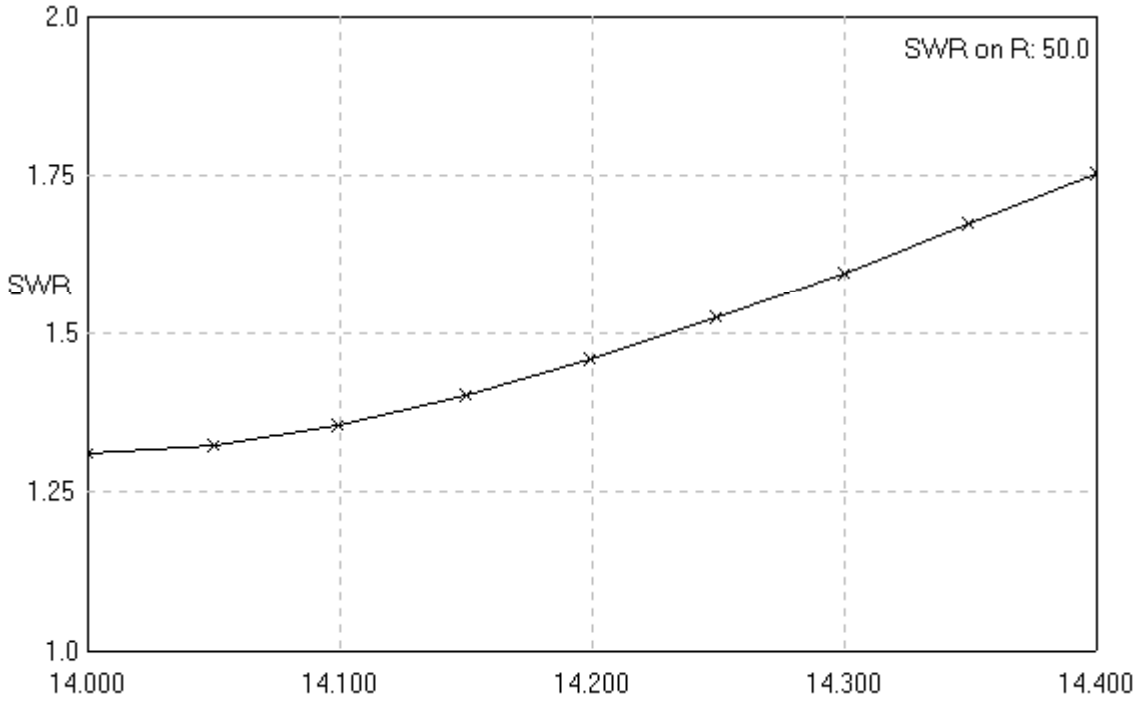


Diagram Directivity and Input Impedance at 20 meters for 4B- W3DZZ Antenna

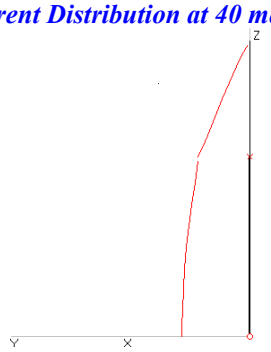


Freq	R	jX	Ga	F/B	ON
14.000	38.9	4.7	-0.2	0.0	On
14.100	40.9	10.5	-0.2	0.0	On
14.200	42.9	16.2	-0.2	0.0	On
14.300	45.1	21.9	-0.2	0.0	On
14.400	47.3	27.6	-0.2	0.0	On

SWR at 20 meters FOR 4B- W3DZZ Antenna



Current Distribution at 40 meters



Current Distribution at 20 meters

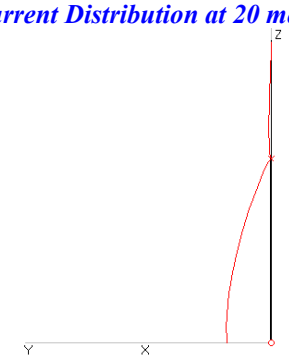
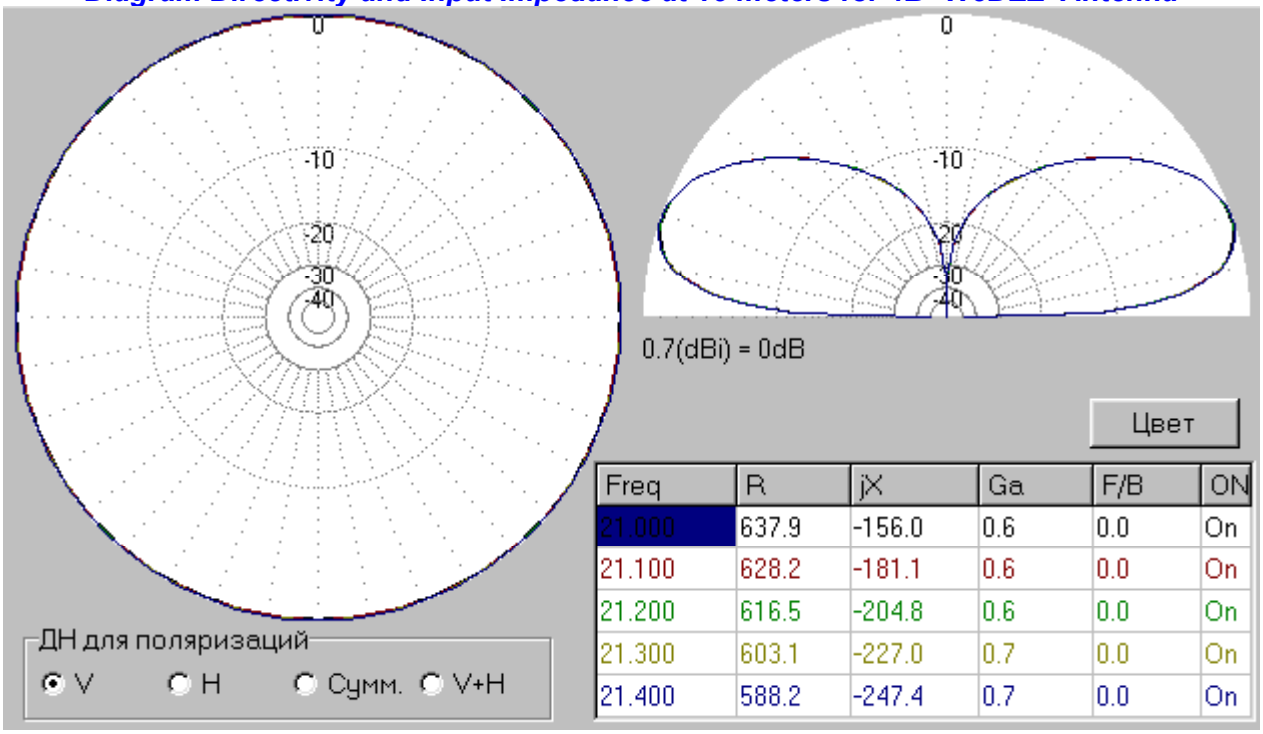
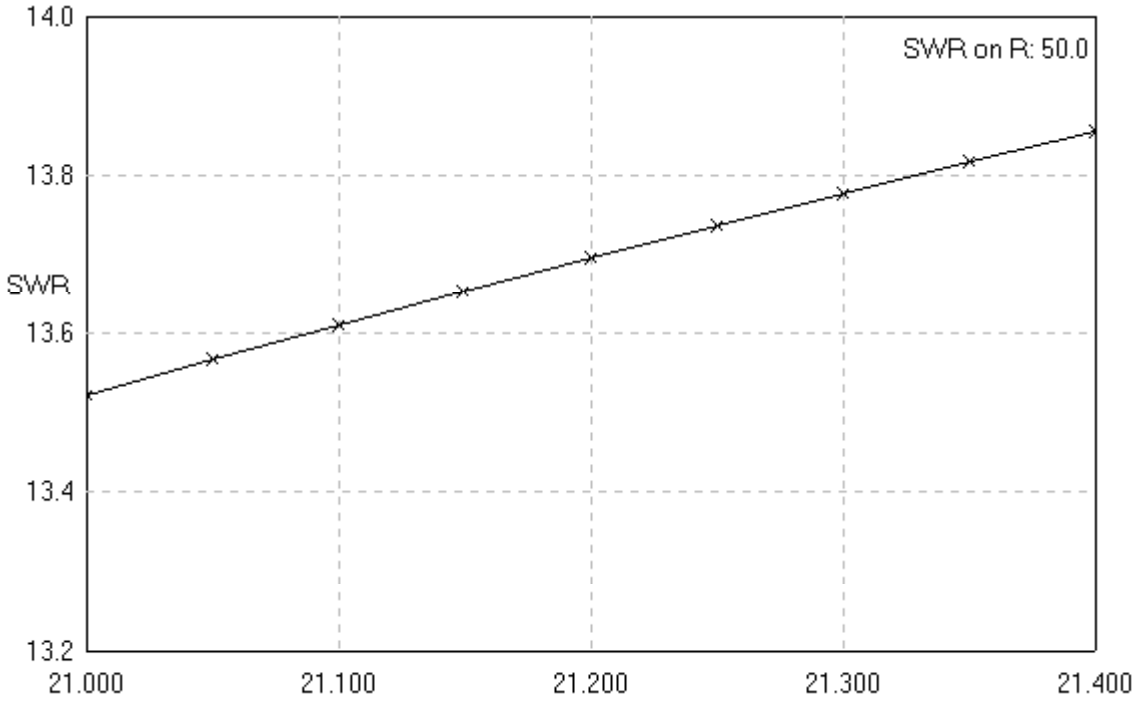


Diagram Directivity and Input Impedance at 15 meters for 4B- W3DZZ Antenna

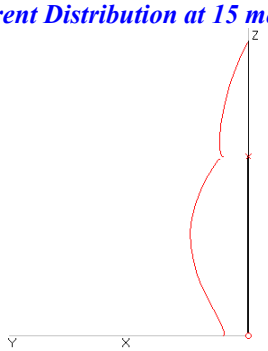


Freq	R	jX	Ga	F/B	ON
21.000	637.9	-156.0	0.6	0.0	On
21.100	628.2	-181.1	0.6	0.0	On
21.200	616.5	-204.8	0.6	0.0	On
21.300	603.1	-227.0	0.7	0.0	On
21.400	588.2	-247.4	0.7	0.0	On

SWR at 15 meters FOR 4B- W3DZZ Antenna



Current Distribution at 15 meters



Current Distribution at 10 meters

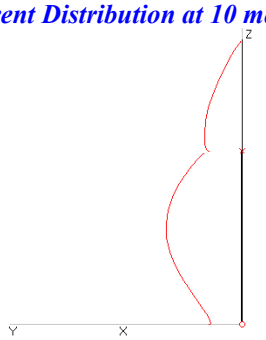
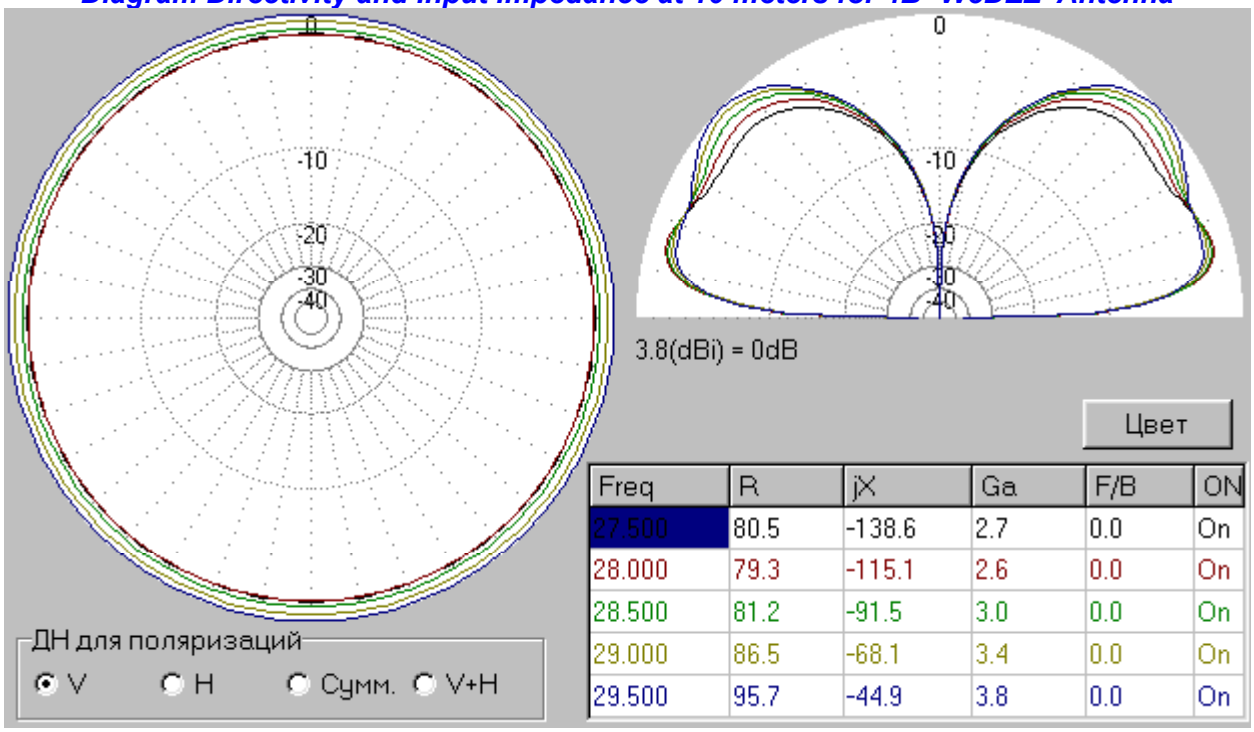
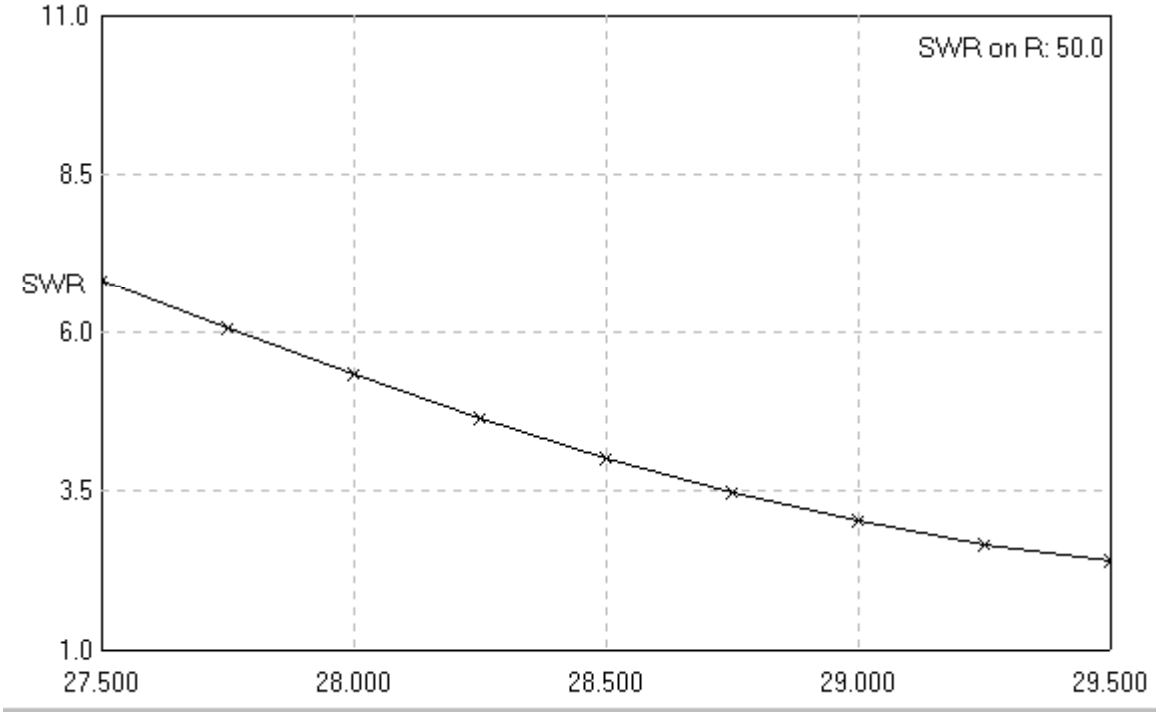


Diagram Directivity and Input Impedance at 10 meters for 4B- W3DZZ Antenna



Freq	R	jX	Ga	F/B	ON
27.500	80.5	-138.6	2.7	0.0	On
28.000	79.3	-115.1	2.6	0.0	On
28.500	81.2	-91.5	3.0	0.0	On
29.000	86.5	-68.1	3.4	0.0	On
29.500	95.7	-44.9	3.8	0.0	On

SWR at 10 meters FOR 4B- W3DZZ Antenna



Trap design: Trap spool contains 10 turns of copper wire diameter in diameter of 2-mm, form of the spool has diameter of 60-mm, distance between turn is 4 mm. The LC circuit should be tuned to frequency of 14.2-MHz. It is possible use a GDO for the tuning. A capacitor at 3-pF is bridged to trap capacitor when the trap is tuning to the resonance. The capacitor is simulated a stray capacitance of the antenna sections. It is necessary to safe trap capacitor from the atmospheric effect.

Antenna tuning: At first, with the help of a GDO tune trap to 14.2-MHz. The circuit tune separately from antenna. At second, tune length A to a minimum SWR in 20 meters. At third tune length of the Section B to minimum SWR at 40 meters.

Common notice for vertical multi range trap antennas

Counterpoises: For a proper work a vertical trap

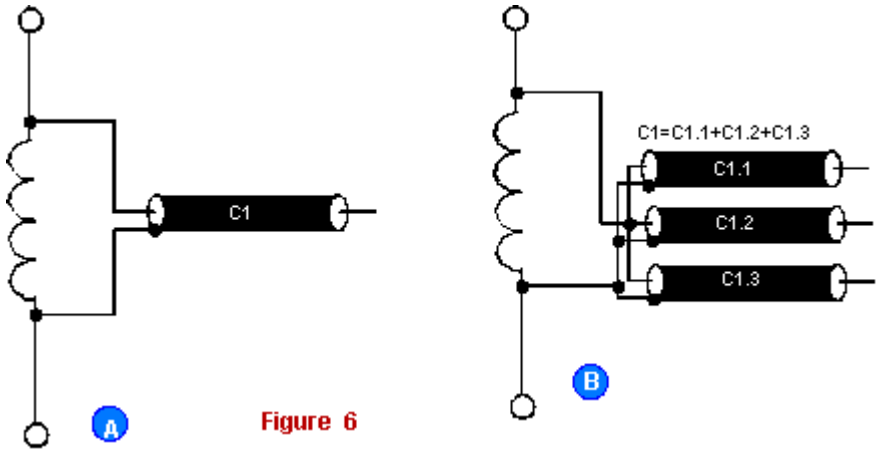


Figure 6

Figure 6 *A coaxial cable capacitor*

antenna need resonance (a quarter wave) counterpoises for each operation range. Use not less than two counterpoises. If the antenna is placed at a small altitude above a metal roof and braiding of feeding coaxial cable have good electrical contact with the metal roof, the antenna can be used without any counterpoises.

Antenna feeding: Quite possible to use a 50- Ohm coaxial cable for vertical trap antenna feeding. Also it is possible to use a two wire line for dipole and vertical trap antenna feeding. In this case it need ATU between the line and the transceiver.

A trap capacitor: A high voltage is at a trap capacitor when the antenna works to transmission. So it need a high voltage capacitor trap to be used at a trap. Such capacitor is costly and rather rare. Hams often use a length of a coaxial cable instead of a high-voltage capacitor. A 50- 0hm coax has near 100-pF/meter, a 75-0hm coax has near 70-pF/meter. Coaxial cable capacity can be find off from a data sheet for the coaxial cable or is metered practically. **Figure 6** shows a coaxial cable capacitor. For a capacitor with a small capacity (up to 30-pF) it is possible to use whole coaxial cable length, see **Figure 6A**. For a capacitor with a high capacity cut the coaxial cable on to several lengths, as it shown in **Figure 6B.**As a high-voltage capacitor in trap it is possible to use a bilateral PC-board by width of 1 to 3 millimeters. In this case capacitor get more bulky the made on coaxial cable basis. It is possible to tune the PC – capacitor on necessary capacity by slitting a foil on one of two sides of this capacitor. Do not forget about atmospheric protection of the trap capacitor.

Other way for trap design: It seems to me in the end of 70s in different radio amateur literature were appeared articles about using “coaxial cable trap” for W3DZZ. There is very simple method for trap making. **Figure 7** shows the trap. It is wise way for trap design, but radio amateur should have an experience using the method. In different radio amateur literature there are a lot of data for design of the trap, but classical methods are described at reference **[6]**.

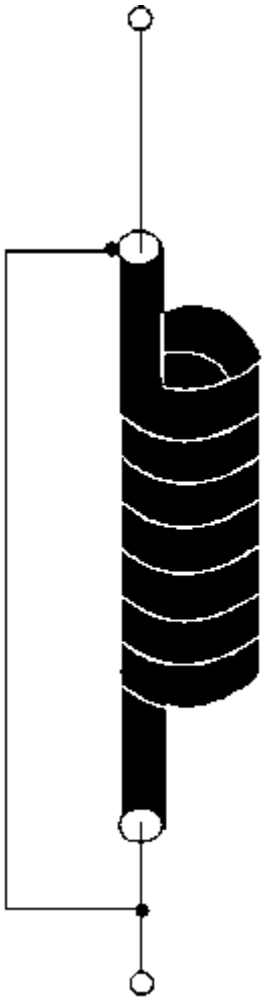


Figure 7

Figure 7 *A coaxial cable trap*

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4. Buchman C. L., W3DZZ : The multimatch Antenna System. // QST, March 1955, pp.22-23, 130.
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A Five Bands Vertical Trap Antenna

M. Chirkov, UL7GCC

The classical **W3DZZ** antenna in vertical installation designed by UL7GCC and shown at the **Reference 1** is well known in Russia. **Figure 1** shows the antenna. Diameter of sections A and B is 40- 50-mm. How is it work?

40-m band: The trap LC cut out the upper section B from the antenna. So only section A works as a radiator, and the section A has length in 10.1 meters, i.e. has electrical length in 1/4λ. Vertical radiator having with the length of 1/4λ has a quarter- wave resonance and works in very effectively way. At the band the circuit LC works as a **trap**.

80-m band: On the 80-m band the antenna has summary physical length of this two sections A+B a little less than 1/4λ. A + B = 16.47 meters, less then 20 meters OF quarter wave length for the 80-m band. A short vertical radiator has a capacity part in its input impedance. But the circuit LC at the 80-m range has an inductance part in its impedance. The inductance part compensates capacity part of the electrically short vertical, and the antenna has a low SWR at the 80-m range too. In other words, the inductor of the LC works as a usual lengthening spool.

10-, 15- and 20-m ranges: Visa versa**,** at the 10-, 15-, and 20-m the LC has a capacity part at its impedance that goes the electrical length of the antenna to 1.75λ at 10-m, to 1.25λ at 15-m and to 0.75λ at 20-m.

Do not forget, verticals like counterpoises, so use several 1/4λ counterpoises for each bands.

Reference

M. Chirkov, UL7GCC: Multi range vertical // Radio #12, 1991, p. 21.

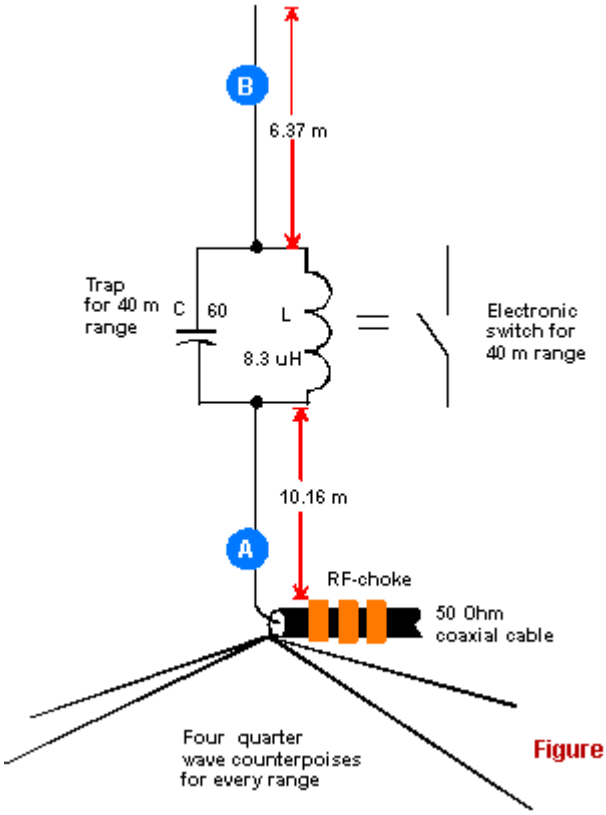

Figure 1

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DIPOLE NADENENKO

Soviet radio amateurs well know the broadband dipole named in Russia "dipole Nadednenko." The antenna is widely used at serve radio centers of Russia. Russian radio amateurs also are used the dipole. Below we take up a design of the antenna.

The dipole contains several wires at each shoulders shaped as a cylinder. Figure 1 shows the dipole Nadednenko. For working at 40- 10 meters the sizes are: L= 8 meters, L1= 3 meters, L2= 1 meter, 2R= 1 meter. Diameter of wires is 1.5- 3 millimeters.

Dipole struts can be both as metal as wooden. As usual, struts has the shape as a circle, wooden struts has the s polygon. Wires are attached to struts any possible way. ' ends of shoulders carefully are welded. As usual, a 300- wire line is used for feeding of the antenna. Antenna rad with horizon polarization.

By Radio 1959

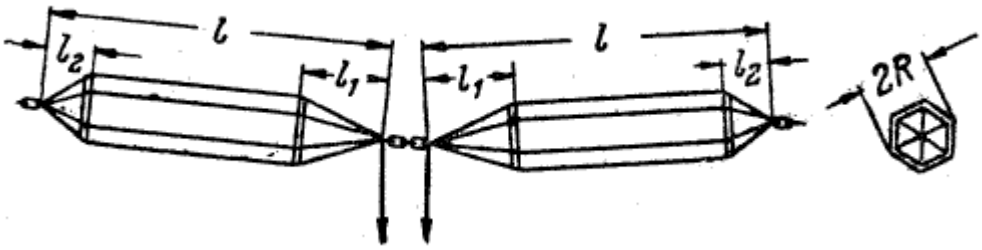
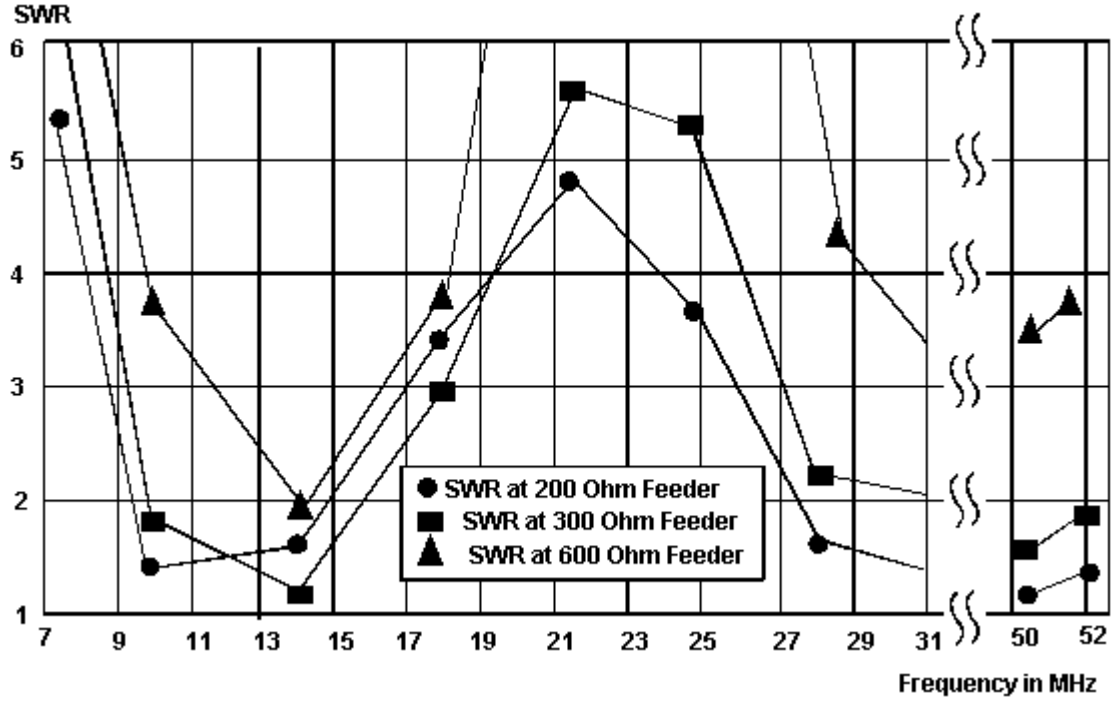
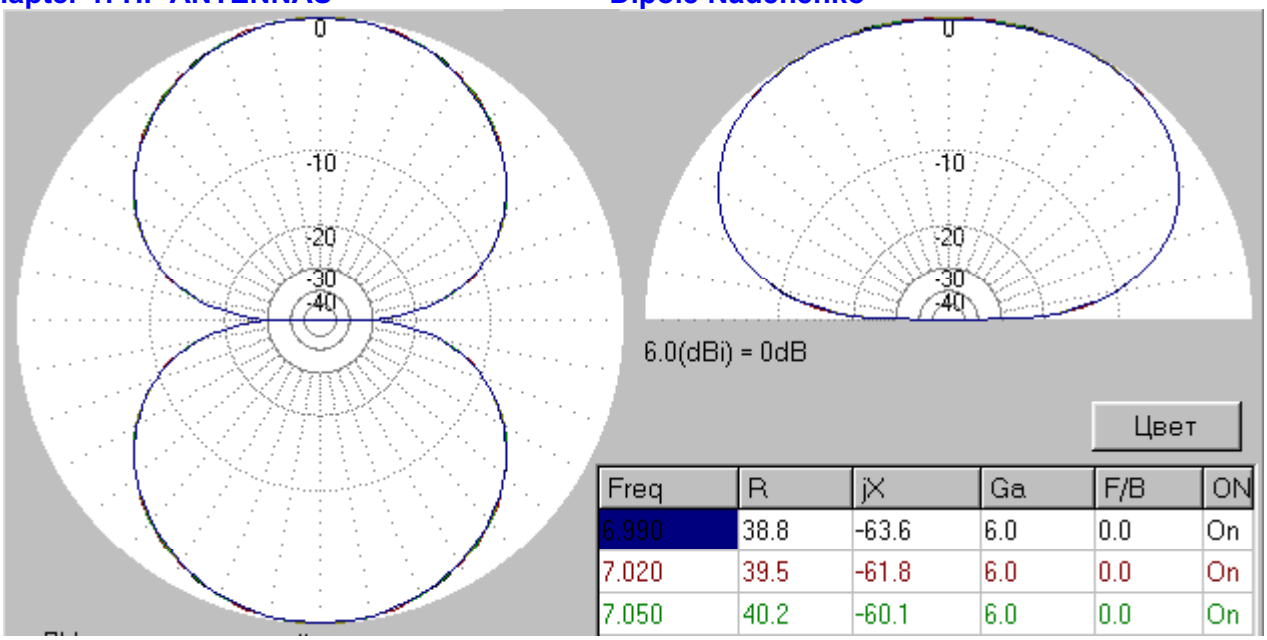


Figure 1

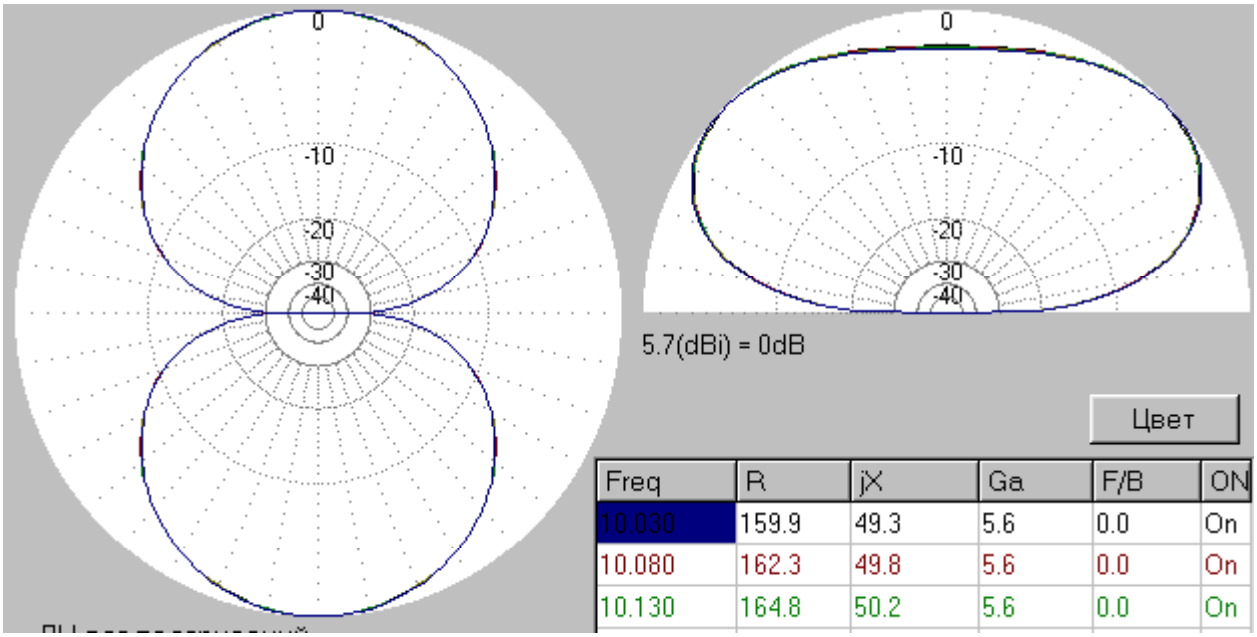
Calculations of input impedance and DD (***for horizon radiation***) of the dipole Nadenenko located at 10 meters above real ground with above mention dimensions (L= 8 meters, L1= 3 meters, L2= 1 meter, 2R= 1 meter, diameter of wires is 2 millimeters) are shown below. You can see, it is possible to use a 50- Ohm coaxial cable with a 1:4 transformer if restricted bands (30, 20, 10 and 6 meters) are used.

The data is obtained with help of a free antenna program MMANA (MININEC based). Left diagram is a section of the volumetric diagram directivity of plane X- Y at a zenith corner of the maximum radiation. The right diagram is section of the volumetric diagram directivity of plane X- Z. Also at the right down corner of the pictures is a table with antenna impedance.

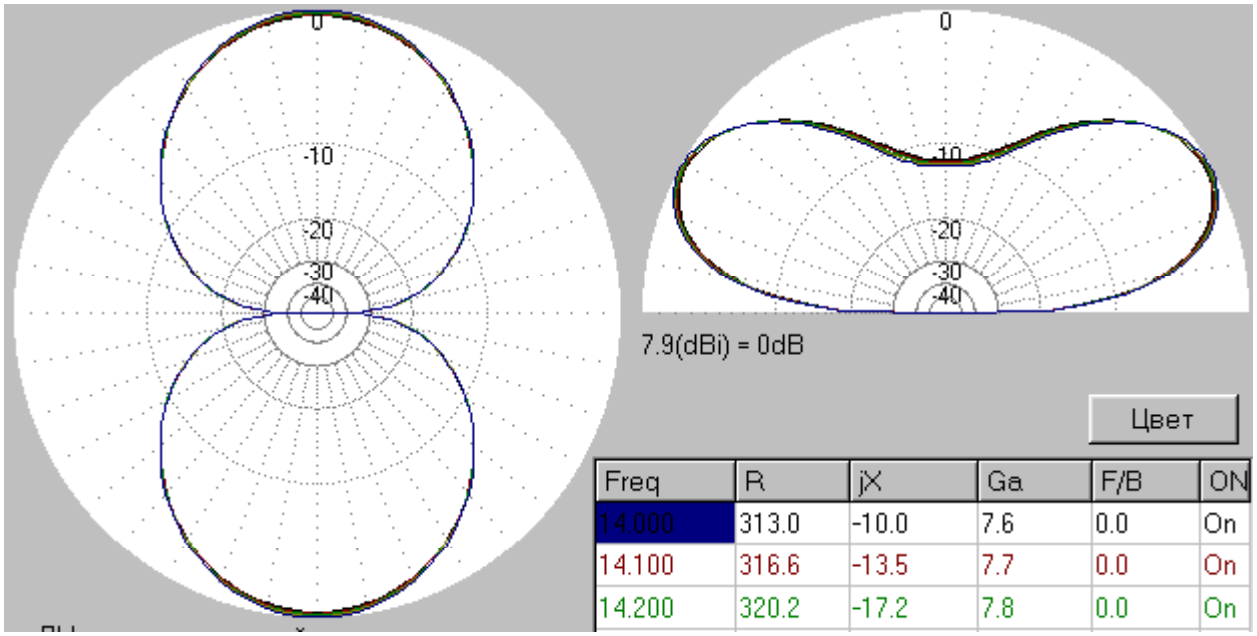




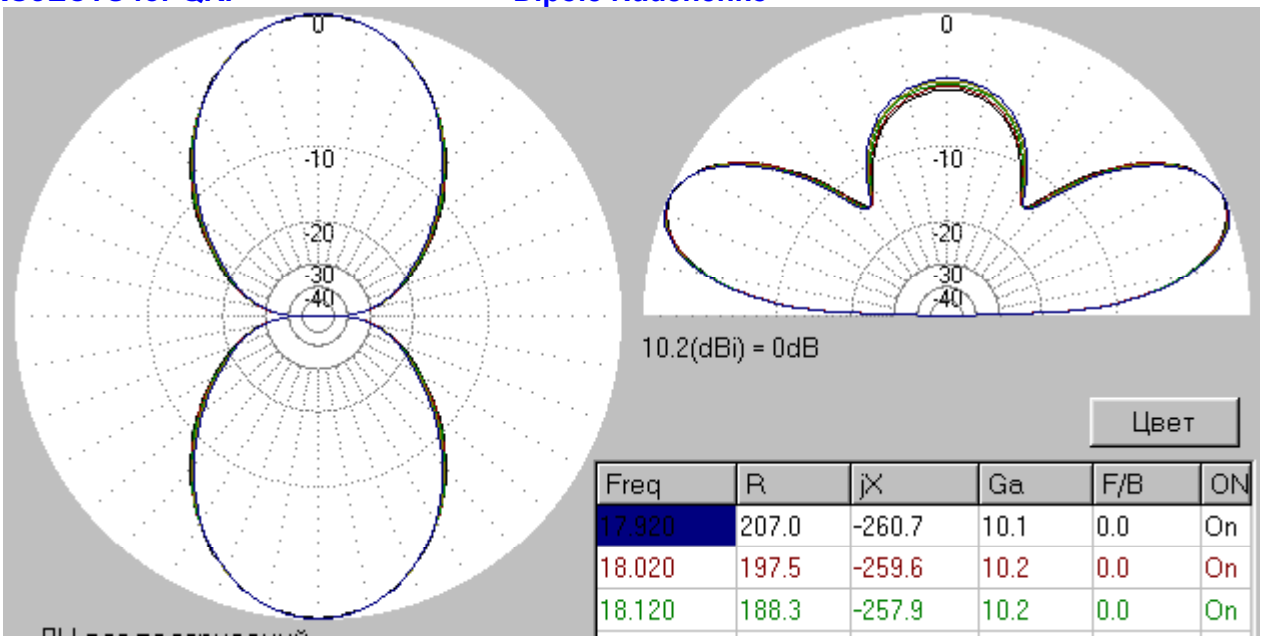
Freq	R	jX	Ga	F/B	ON
6.990	38.8	-63.6	6.0	0.0	On
7.020	39.5	-61.8	6.0	0.0	On
7.050	40.2	-60.1	6.0	0.0	On



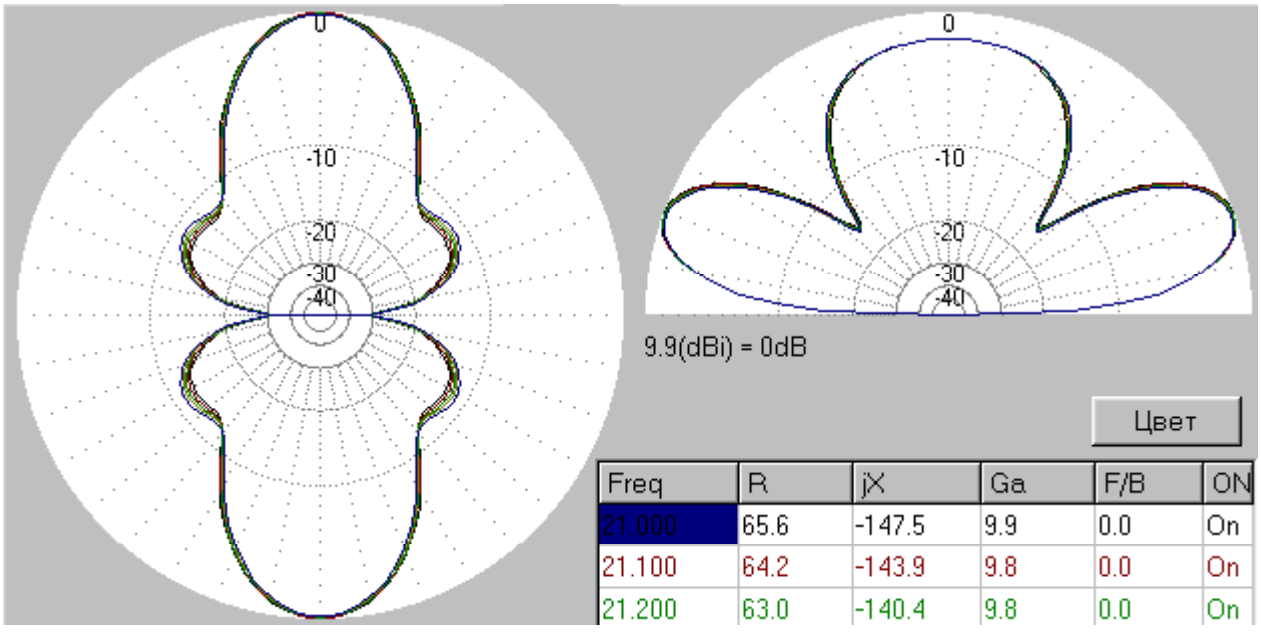
Freq	R	jX	Ga	F/B	ON
10.030	159.9	49.3	5.6	0.0	On
10.080	162.3	49.8	5.6	0.0	On
10.130	164.8	50.2	5.6	0.0	On



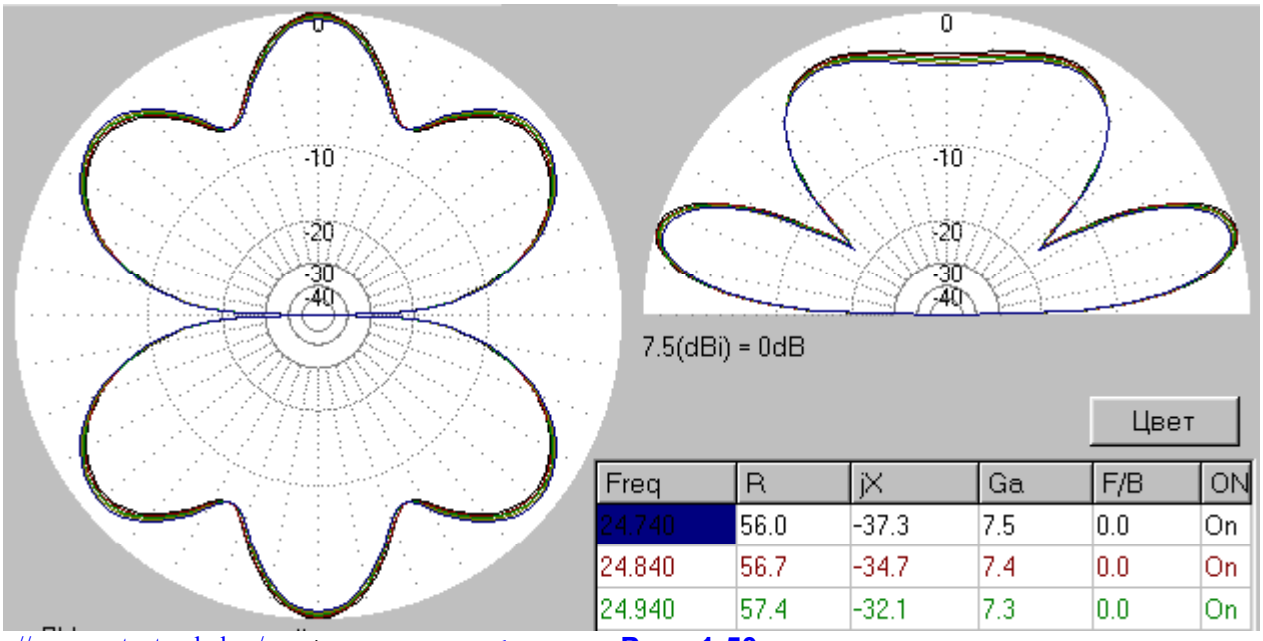
Freq	R	jX	Ga	F/B	ON
14.000	313.0	-10.0	7.6	0.0	On
14.100	316.6	-13.5	7.7	0.0	On
14.200	320.2	-17.2	7.8	0.0	On



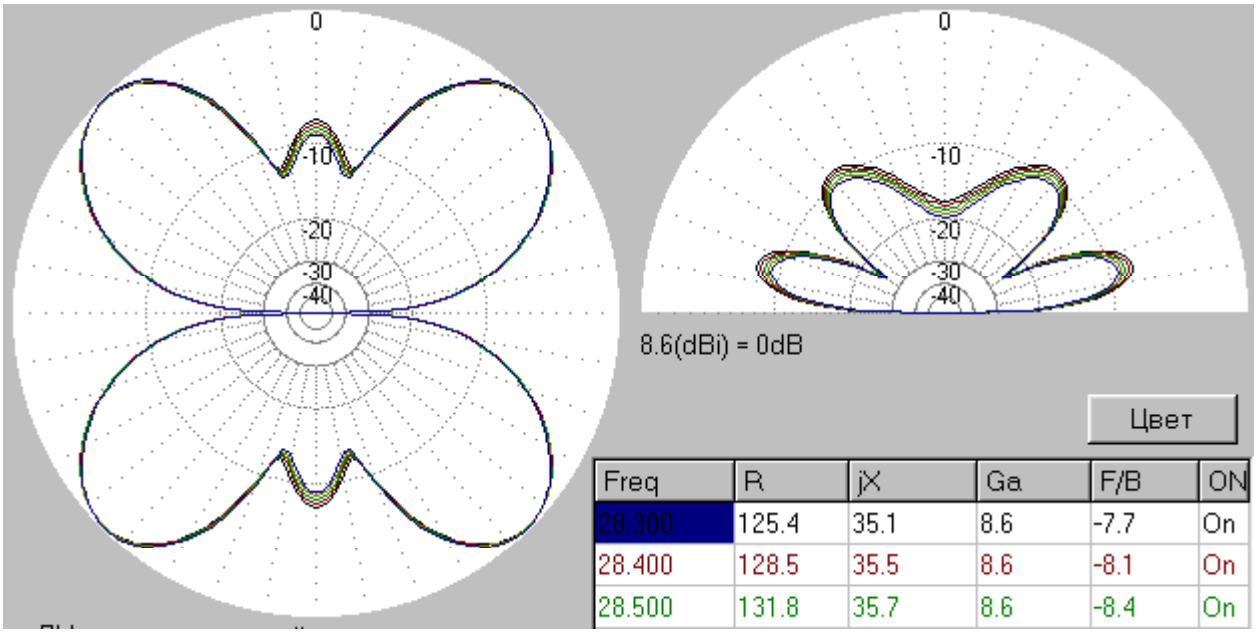
Freq	R	jX	Ga	F/B	ON
17.920	207.0	-260.7	10.1	0.0	On
18.020	197.5	-259.6	10.2	0.0	On
18.120	188.3	-257.9	10.2	0.0	On



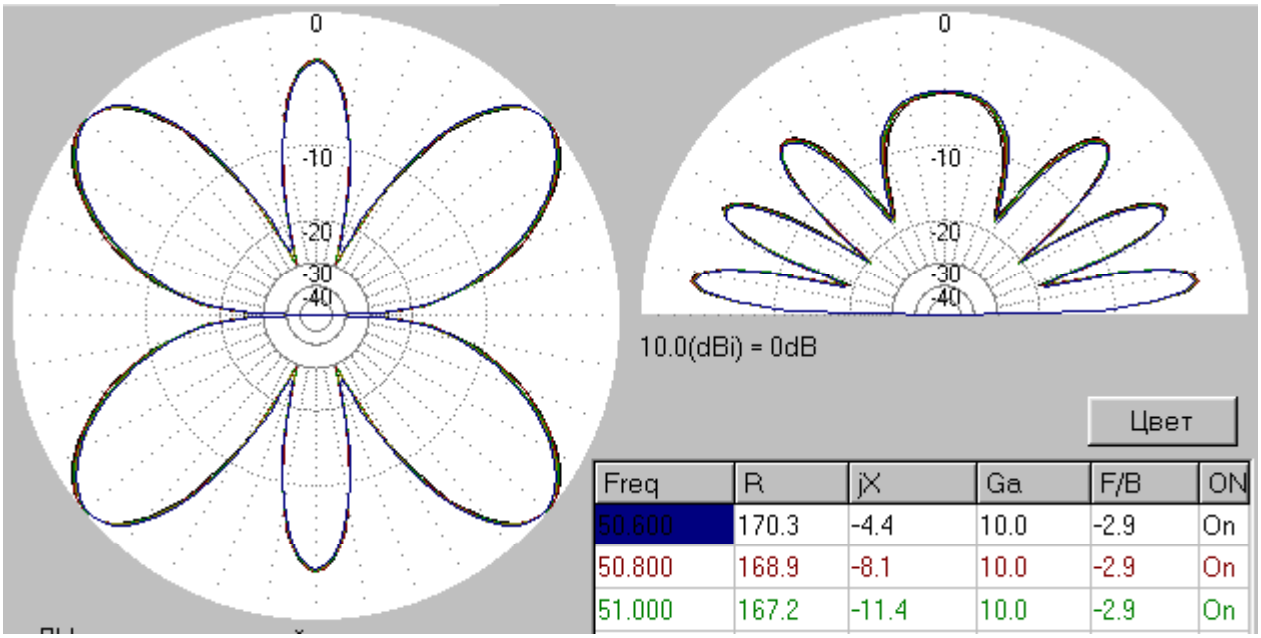
Freq	R	jX	Ga	F/B	ON
21.000	65.6	-147.5	9.9	0.0	On
21.100	64.2	-143.9	9.8	0.0	On
21.200	63.0	-140.4	9.8	0.0	On



Freq	R	jX	Ga	F/B	ON
24.740	56.0	-37.3	7.5	0.0	On
24.840	56.7	-34.7	7.4	0.0	On
24.940	57.4	-32.1	7.3	0.0	On



Freq	R	jX	Ga	F/B	ON
28.300	125.4	35.1	8.6	-7.7	On
28.400	128.5	35.5	8.6	-8.1	On
28.500	131.8	35.7	8.6	-8.4	On



Freq	R	jX	Ga	F/B	ON
50.600	170.3	-4.4	10.0	-2.9	On
50.800	168.9	-8.1	10.0	-2.9	On
51.000	167.2	-11.4	10.0	-2.9	On

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Russian Military WW - II Radio "RBM"

Radio RBM is one of the most famous Russian military radio that was used in the WW-II and after the war as a surplus radio. RBM took place in the WW-II, after the war RBM was used as trial radio for military teaching centers. Lots of Russian hams know well the radio. I want to give some information about RBM and its antennas at ANTENTOP.

73! I.G.

History: Radio RBM, firstly named as RB (Radio, Base), was designed before the WW-II, in the 1938, in the Research Center of Communication of Red Army, by a special research group guided by colonel Sosunov. Special variant of RB named as RB-40, that had low weight because it was made in an aluminum cabinet was produced from end of 1939 for spy and partisan. It was made near 1000 RB-40, but then its producing was stopped because the aviation need aluminum. In the 1942 some modifications were done in the radio, and RB was named as RBM (Radio, Base, Modified). In the 1943 both with RBM was produced RBM-5 that has 5 watts power compare to 1 watt that RBM has. After WW-II other modification of RBM named as RBM- 1 was produced. RBM-1 was produced until end of 50s. Some samples of RBM-1 was produced for export (***see picture with Latin letters on the front panel***).

Composition: RBM –1 consists of from two boxes, one is the transceiver other is its supply unit.

Transceiver has dimension of 345x195x260-mm, weight of 13 KG. It works at two frequency ranges, I- 5.0- 2.75 MHz, II - 2.75-1.5 MHZ, has CW and AM modes. Transmitter made on vacuum tubes

Transmitter has 1 watt output(in reality 1.5 watts), plate current 35-mA and heater current 1-A.

Receiver has sensitivity of 10-μV at AM and 3-μV at CW, plate current 10-mA and heater current 0.5-A.
Transmitter and receiver use common units as: an antenna and output audio transformer, that does modulation for transmitter at AM mode and audio at receiving mode.

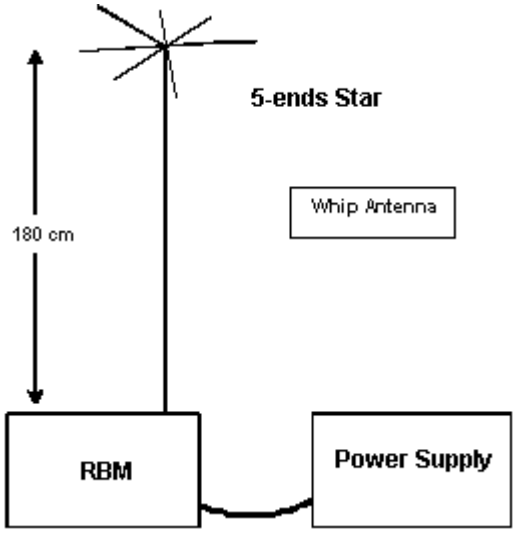
Power supply has three batteries of BAS-80 for plate and a NiCad accumulator 2NCN-24 for heater, weight of 14 KG. The Power Supply run the radio during 24- 36 hours.

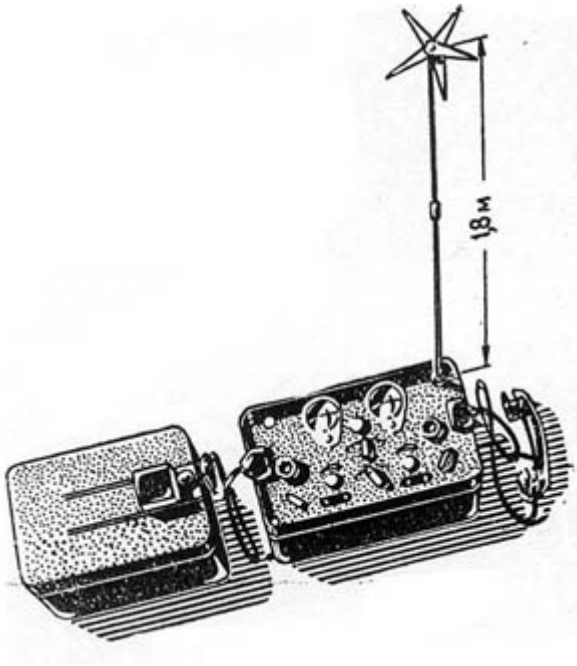
Purpose of RBM is to do reliable simplex communication at any conditions. Distance of the communication depends on antennas that use with the radio.

WW- II. USSR, Leningrad Front, 1943. Radio RB

Export Sample of RBM-1

Short Whip Antenna does communication near 10 kms on AM and near 15 kms on CW.



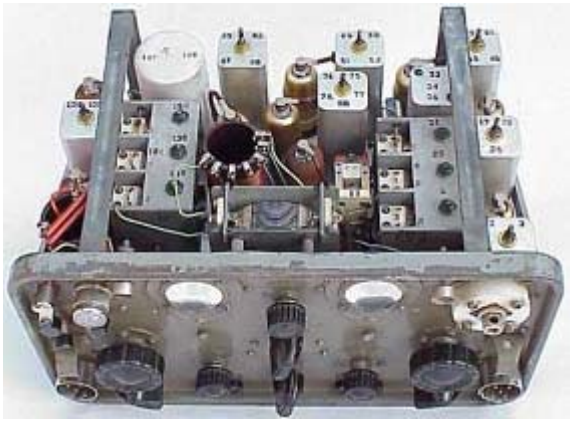


I live near village Prohorovka, Russia, where at July-4- August 5, 1943, The greatest tank battle of WW - II was. As I know, near 500 samples of RB radio took place in the battle

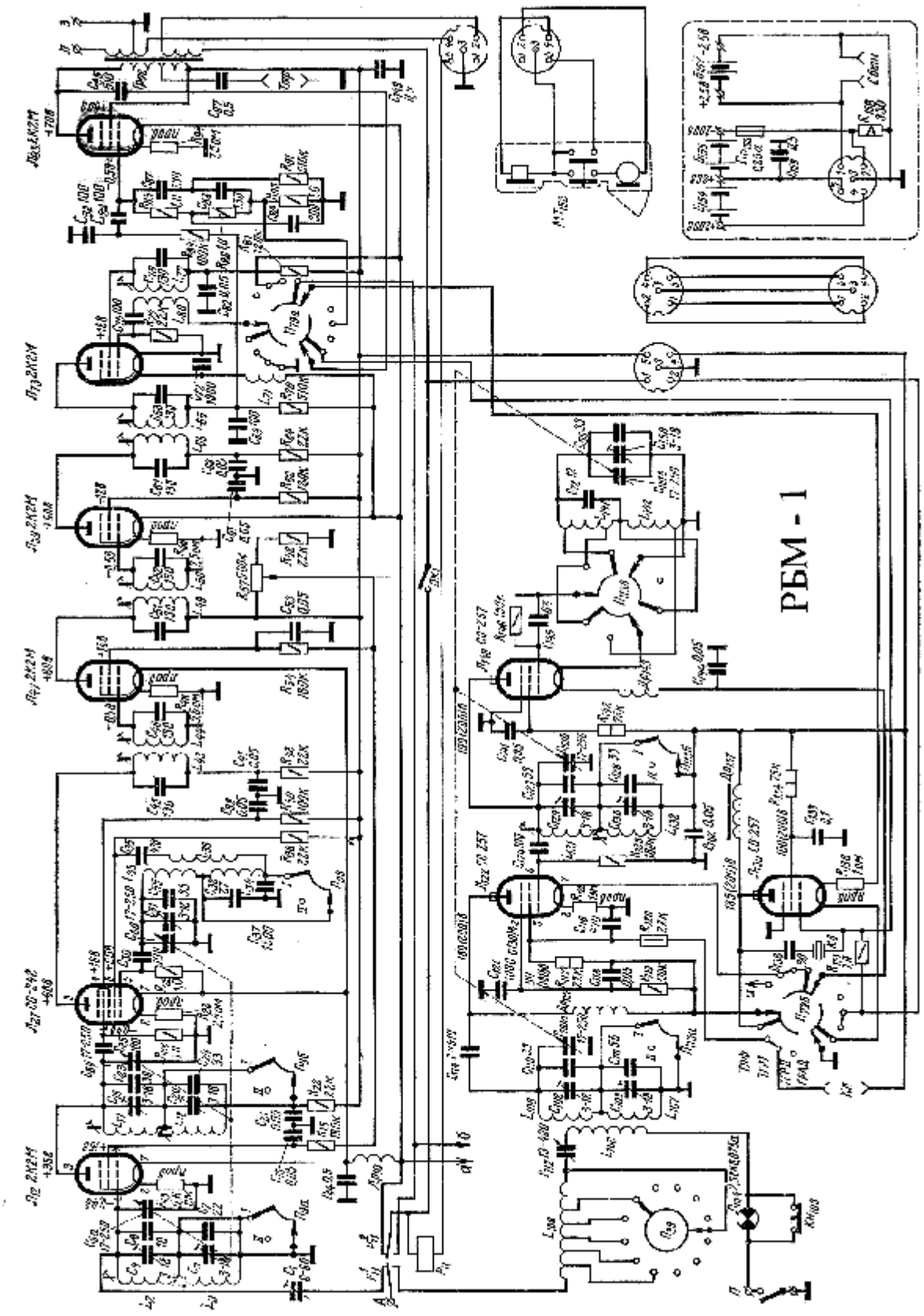
German tank Elefand/Ferdinand
Destroyed by a tank mine.
Prohorovka, July, 1943

USSR tank KV-1
Destroyed by a German tank Tiger.
Prohorovka, July, 1943

Russian RBM-1



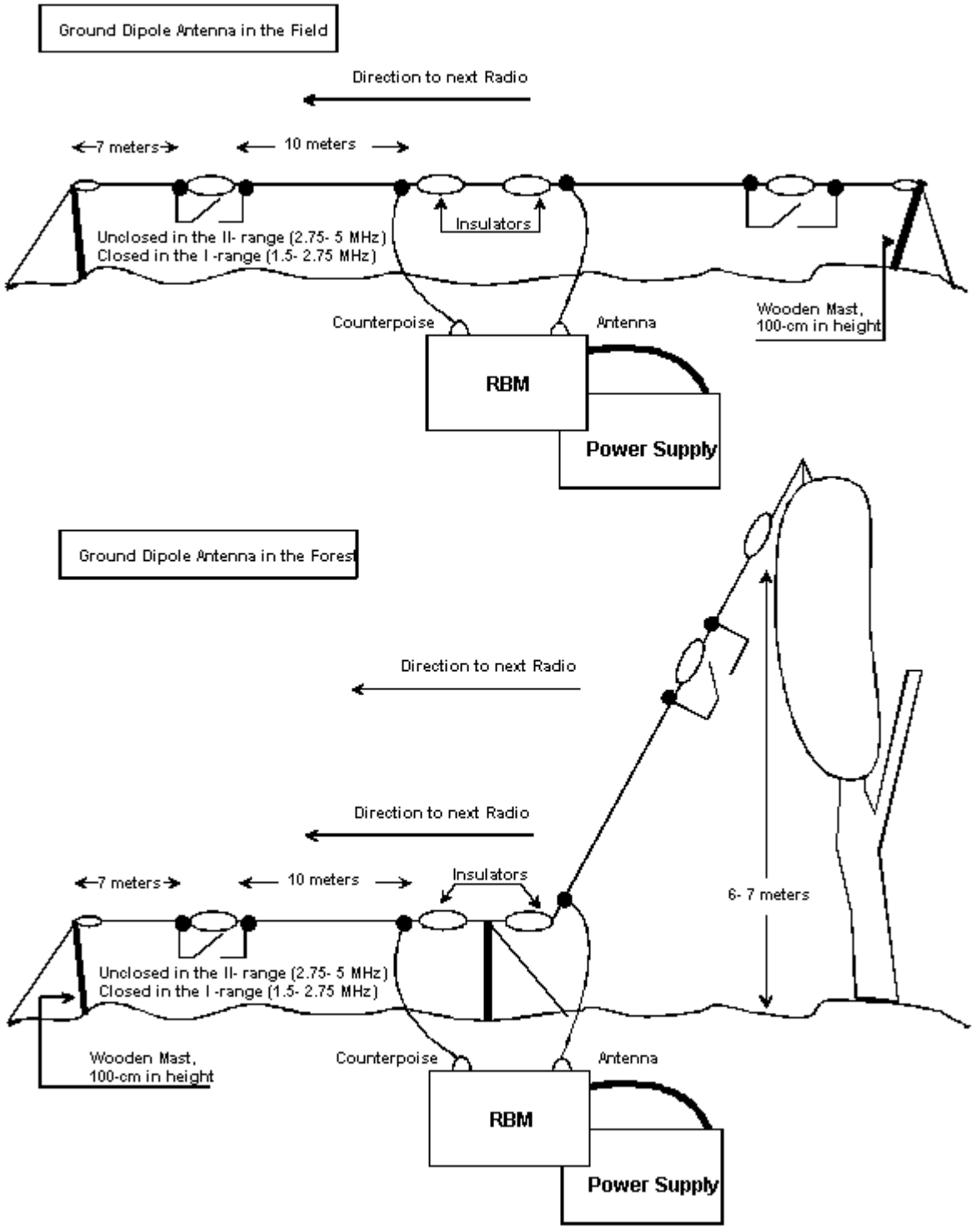
Schematic of RBM-1



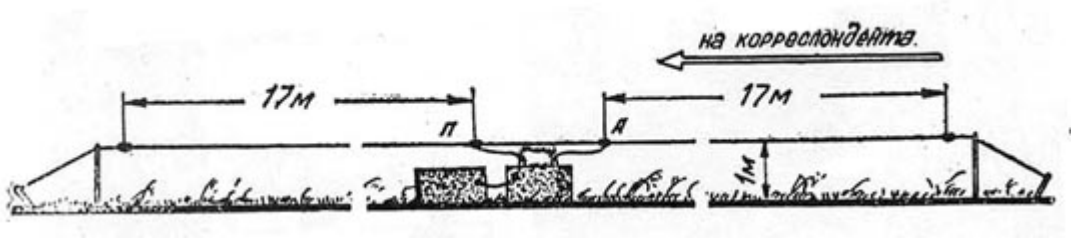
Ground Dipole Antenna does communication near 17 kms on AM and near 35 kms on CW.

Mast Antenna does communication near 30 kms on AM and near 50 kms on CW.

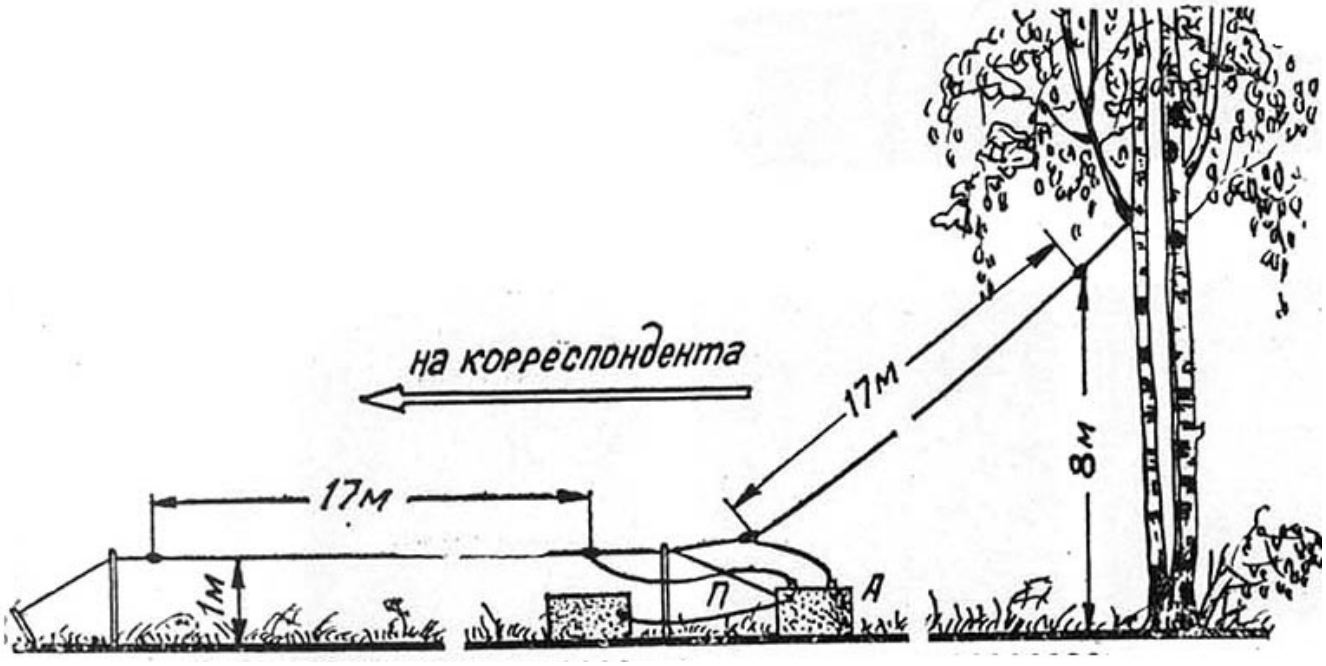
Note: The distance of communication is shown for daytime at middle level of interferences. At nighttime and at high level of interferences the distance of communication is decreased in two times.



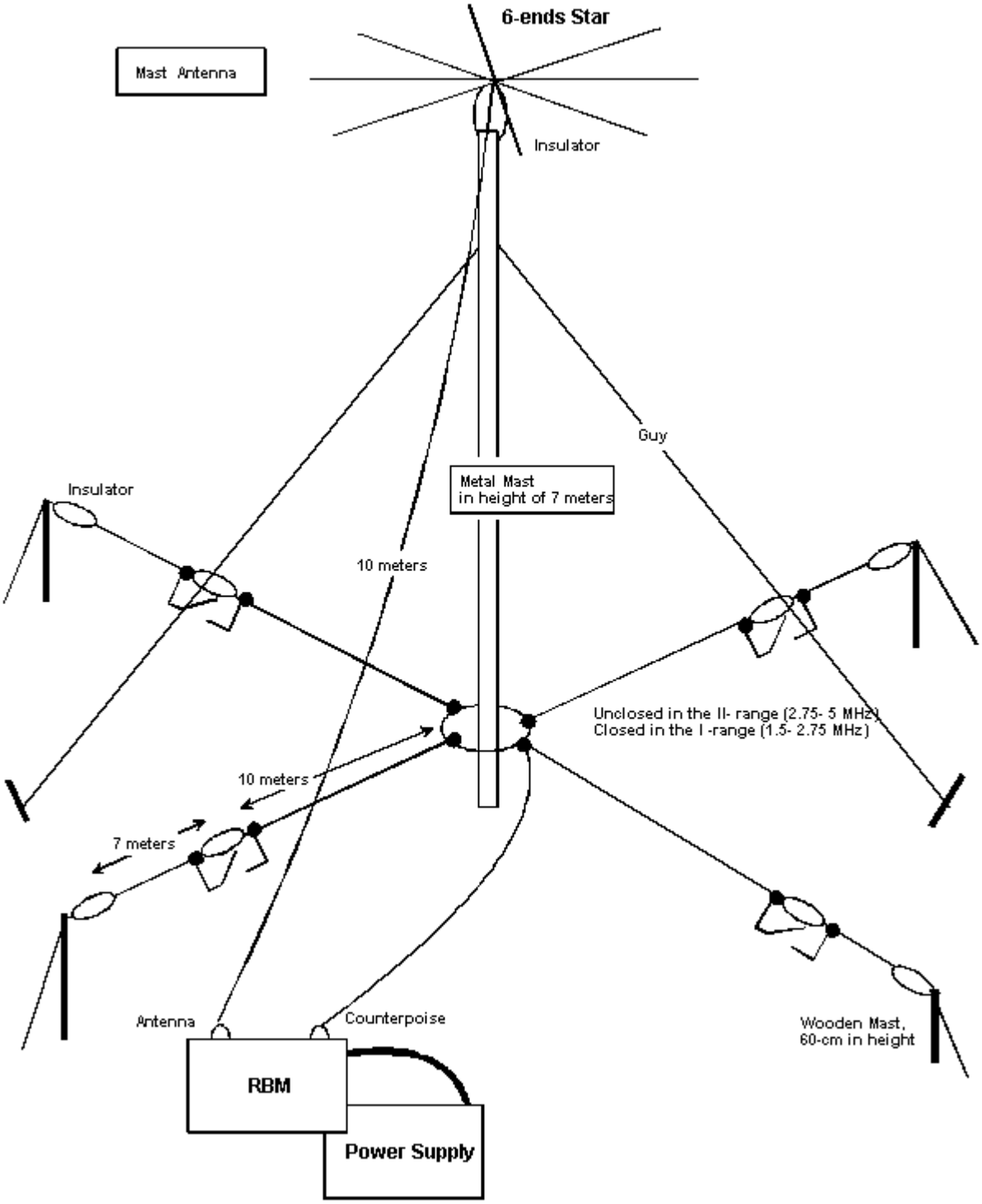
Picture from Russian Manual



Picture from Russian Manual



Winter, 60s, XX- century, Russia



Credit Line:

Radio Magazine, USSR

RBM Padio. Manual for user. 1952.

http://www.cqham.ru/

http://www.oldradio.onego.ru/

http://www.battlefield.ru/

HALF - LOOP ANTENNAS
FOR HF COMMUNICATIONS IN ALE AND FREQUENCY HOPING

Jean-Pierre GOUIN & Daniel LAFARGUE

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The present describes a HF loop antenna and its agile coupler which can be adapted to the new designs of ALE and FH (frequency hopping) radiosets. The original specification in 1993 was : "a small mobile antenna and coupler for HF voice and data communications in driving from 0 to 600 km without silent zone, in association with a 125 Watts CW radioset.

Frequency range 2-12 MHz Channel tuning time < 5ms Bandwidth > 3,5 kHz in a military environment" Following on from this product other versions with wider frequency range (3-15 MHz, 3-30 MHz, 2-30 MHz), a higher power and various dimensions and shapes for fixed, land-mobile and naval applications have been developed.

1. GENERALITIES

1.1. on the HF tuned loops

The HF transmission tuned loop antennas which are designed for HF transmission have small dimensions (< 0,1λ) compared to the wavelength, in order to conduct a quasi constant current and to be considered as magnetic dipoles. Their radiation impedance and efficiency mainly depends on their surface which creates a magnetic flux in the near field and an electromagnetic field in the far field. Their diameter, height or width (round or square shape) run from 1 to 3 meters, and their radiating surface generally do not exceed 5 m^2 in order to coincide with the small dimensions required.

These types of antennas differ from open antennas (like whips, horizontal dipoles, log-periodic antennas,) by their impedance which is reactive and can be adapted by capacitor only. Their radiating resistance is low (< 1 mΩ) at the lowest frequencies of the range. As the efficiency is given by the ratio radiating resistance/ total resistances of the tuned circuit, it is necessary to minimise the radiating element resistor, using a good conductive metal (aluminium, copper), and to use low loss capacitors.

These conditions beeing fulfilled, the loop antennas deliver a high current and have a high selectivity with a high quality factor (Q-factor), typically 10 times higher than the Q-factor of the best traditional couplers designed for 5 to 10 meters whips or 10 to 40 m dipoles.

1.2 On fast frequency tuning

For future fast ALE procedures the tuning target time is 50ms, while the "low speed" frequency hopping (F.H.) procedures already require a 5ms tuning time, with all calculations and control exchange times beeing included or already done. This can't be done using electromechanical tuning. Digital switching devices are cost effective today at low and medium powers. Their switching time run in milliseconds using low loss vacuum relays, and in the microseconds using electronic relays like PIN-diodes. But the PIN-diode technology cannot be used in loops for transmission, due to their inability to withstand the high currents and due to the losses they bring (0.5 to 1 W) which would drastically decrease the overall efficiency at the lowest frequencies. Vacuum relays, including REED relays, are the only technologies available to switch the capacitors of a transmission tuned loop antenna efficiently.

1.3 On the power requirements

Based on the experience of 2 previous generations of tuned loop antennas, and the proprietary propagation simulations, it was calculated that two 100W radiosets and 4m^2 loops having a -15 to + 5dBi typical gain figure from 2 to 12 MHz would insure voice and data communications at any distance from 0 to to 600 km at least.

This mission cannot be fulfiled by any 5 to 10 m whip antenna on a medium soil, even in association with a 400W/1 kW radio set: a vertical whip or a bent whip on a vehicle in move do not transmit and receive enough energy to cover the typical 50-250 km silent zone.

A 125W radioset combined with a tuned loop antenna is sufficient to fulfill the mission requirement using the Near Vertical Incident Signal (NVIS propagation). This will be further improved due to frequency management and the new generations of HF modems which will bring a lower threshold of sensibility.

II DEVELOPMENT OF A NEW MOBILE TUNED FRAME ANTENNA

II.1 Principle

The mobile tuned loop antenna is a "half-loop" set-up vertically on a metal surface which achieves a full loop equivalence. The metal surface like a mobile platform (truck or shelter, ship's cabin,...) must have a good electrical continuity. Thie half-loop is half the size of a full loop and makes installation possible on small vehicles on the move.

The half-loop is folded and joined at each end to the platform's earth. One end is loaded by a variable capacitor.

A feed rod ("the feed coil") links the radioset RF access to a precise point of the half-loop. It is equivalent to a fixed reactive element, and the whole system acts as a loss-free autotransformer whose primary circuit can be set to 50 W.

II.2 Modelisation of the antenna

The modelisation purpose is the definition of the electrical circuit and the parameters of the antenna. It is made by the wire methods of moments.

The radiating element is represented by a radiating impedance (Rr, La) with a loss resistance Rp

The tuning capacitor is represented by a serial circuit (C, Rc), C being the capacitor value and Rr its loss resistance.

The 50Ω matching is figured by a loss-free transformer M with a matching ratio K, and a parallel or serial inductance L at the RF input.

Establishment of the equivalent circuit parameters:

-The radiating element (Rr, La) is calculated by an electromagnetism software based upon the method of moments.

- The radiating element loss Rp is determined according to the antenna material and section

- The capacitor's losses Rc are determined through the manufacturer's data

- The matching ratio K is a function of the primary to secondary radiating surface ratio

- The inductance L is a function of the spiral surface comprised between the feed bar and the platform.

Two types of antennas have been compared, type A and type B, differing by the positions of their capacitors.

II.3. Modelisation of the antenna type A

The capacitor is positioned in the secondary of the transformer, at the end of the line (**FIG 1**).

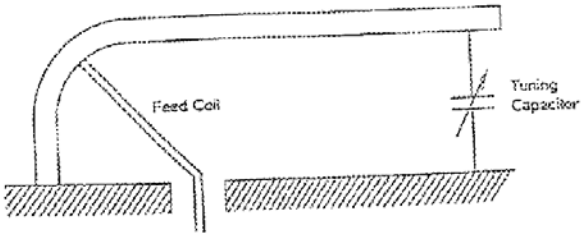


Figure 1

The electrical equivalent sheme is given **FIG 2**

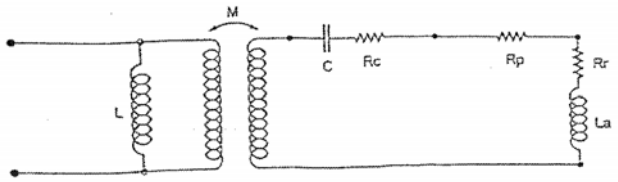


Figure 2

The results are computed by a specific C.A.D. radiofrequency device and compared to the values measured on full scale antenna mock-up.

As an example, **FIG 3**, **FIG 4**, **FIG 5** show the

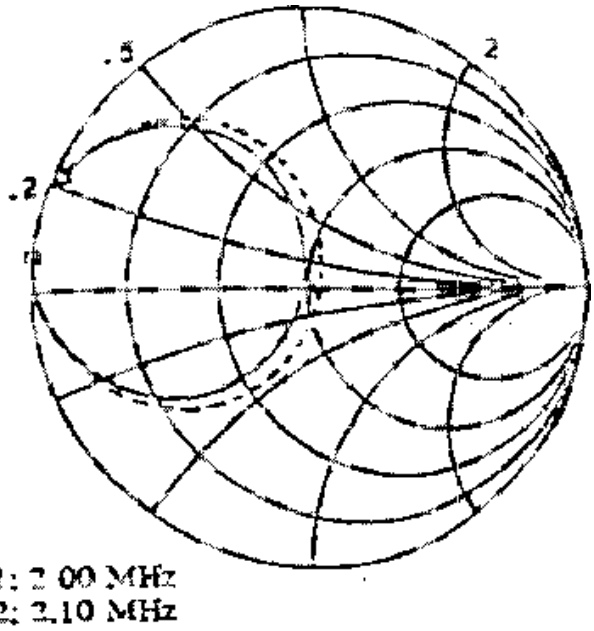


Figure 3

impedances at various frequencies on the Smith charts, with computed values (in full line) and measured values (in doted lines). These charts underscore the performances of a resonating cavity like a R, L, C parallel device, and confirm the impedance values computed by the method of moments.

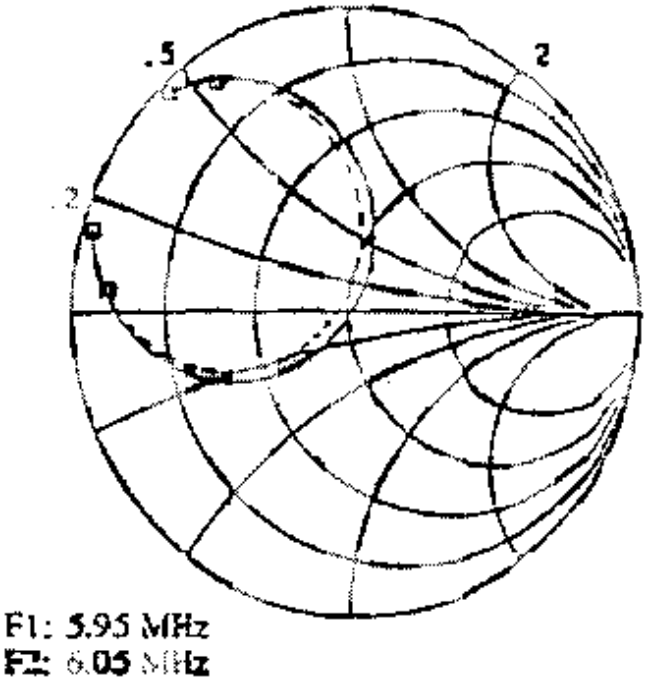


Figure 4

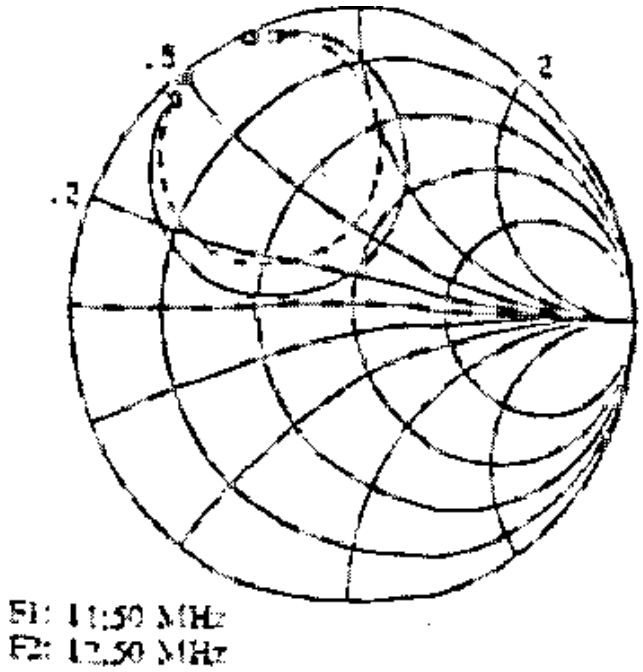


Figure 5

The calculated and measured values are compared at various frequencies (**FIG 6**, **FIG7**) The bandwidth is measured at VSWR ý 2.5:1 , when the real and the imaginary terms of the impedance are equal.

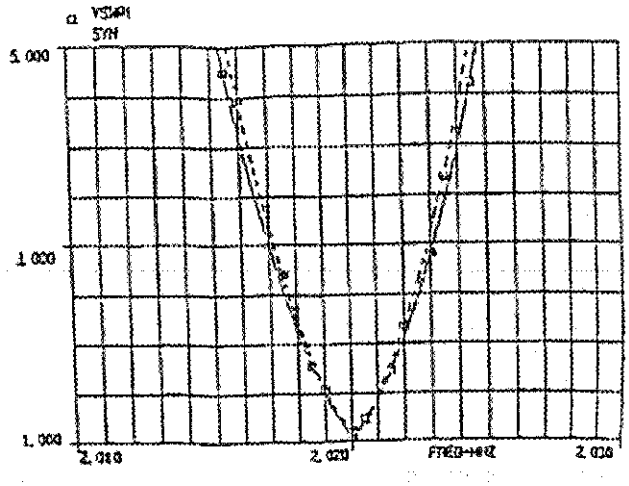


Figure 6

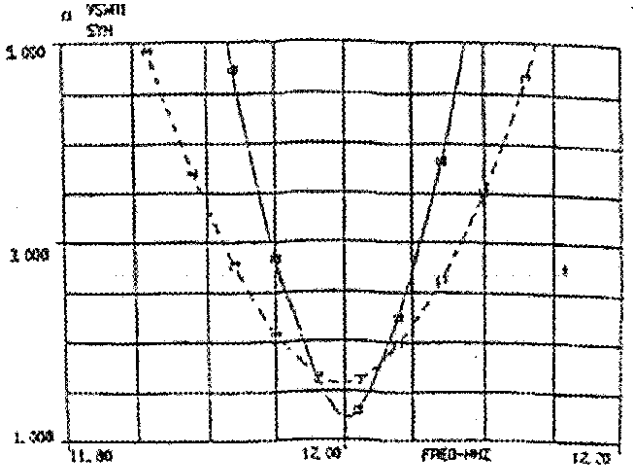


Figure 7

The results have validated the antenna equivalent circuit.

This sheme helped to optimize the dimensions of the radiating element, considering the efficiency and bandwidth requirements. The approximative values are, from 2 to 12 MHz:

Rr = 0,5mΩ to 3 Ωwith a 2,2m^2 antenna surface

Rp= 0.01 to 0.02 Ω

C= 3500 to 60 pF

Rc= 0.05 to 2 Ω

The equivalent circuit aided in the calculation of the voltages and the currents developed over each electronic component.

II. 4. Modelisation of the antenna type B

The tuning capacitor is positioned in the primary of the autotransformer (**FIG 8**).

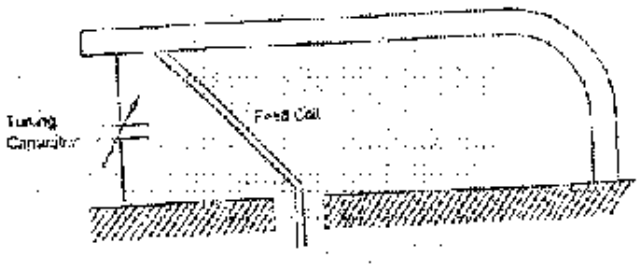


Figure 8

Its equivalent electrical scheme is given on **FIG 9**.

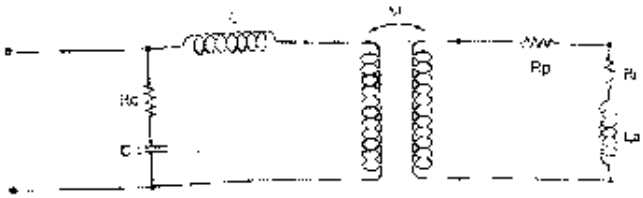


Figure 9

The Type B antenna is modeled in the same way as the Type A antenna, and using the same physical parameters. An additional capacitor may be added in the feed rod to optimize the radioset matching impedance.

II.5. Compared performances type A and type B antennas

Comparative simulations

The compared simulations gave a clear advantage to the Type A antenna type. As an example, **FIG 10** shows a +12 dB gain advantage for the Type A antenna at 12 MHz

Comparative measurements

The comparative simulated results were confirmed by the comparative measured bandwidths Using 2 antennas having the same radiating surface, the compared measured bandwidths were 5 to 10 times

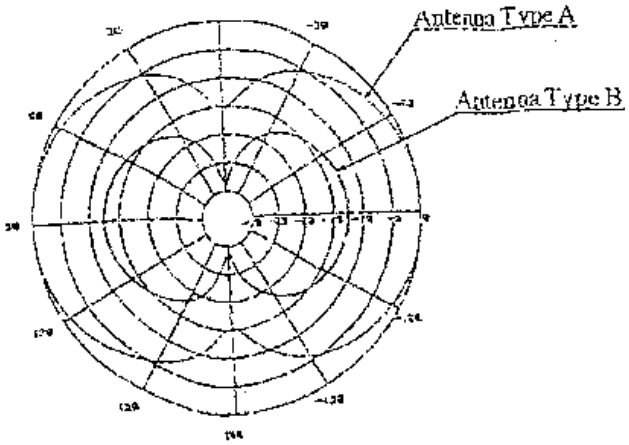


Figure 10

higher for the Type B than for the Type A antenna. In a tuned circuit, bandwidths (B) are inversely proportional to the quality factor (Q), and Q is proportional to the efficiency (h); when Q >> 1, h x B = Rr/2p La = constant If ha and hb are the Type A and Type B antenna efficiencies , and Ba et Bb their bandwidths respectively, the applying formulas are ha Ba = hb Bb, and ha / hb = Bb / Ba When the measured bandwidth ratios is Bb / Ba = 10 , the efficiency ratio becomes ha/hb is 10.

Explanation

Observing that Type B antenna optimizes the tuning in the primary circuit, and that the Q-factors of primary and secondary are quite different, the energy transfer in the secondary is not maximized. On the contrary, in the Type A antenna the tuning brings a maximum Q-factor and the current is the highest in the radiating resistor.

Conclusion

The Type A antenna design brings the best antenna efficiency.

II.6 Improvement of the design

Increasing the bandwith

Trials on vehicles were made under strong rain. Modifications of the tune positions were observed at the highest frequencies (**FIG 11**)

Such modifications can give an operational problem with no possible reset in transmission (in FH mode principally).

This shifting problem was resolved by widening the bandwidth by using two radiating elements in parallel and electrically linked. The simulation of this structure

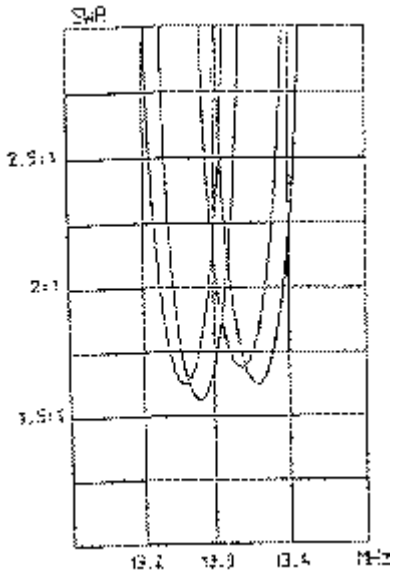


Figure 11

modification concluded in a +10 to +15% extended bandwidths and in +0.5dB to +1dB extra efficiencies all over the frequency range.

II.7. Realisation of a fast tune design

The 2-12 MHz.antenna was developed for the required efficiency and a minimum 3.5 kHz bandwidth independently of the variations in the environment.

With a 2.2m^2 radiating surface the half-loop reactance is 2mH at 2 MHz and 3.5mH at 12 MHz.

The tuning principle consists in switching capacitors in parallel to create a series of bandwidths with mutual covering at a VSWR < 2.5:1.(**FIG 12**)

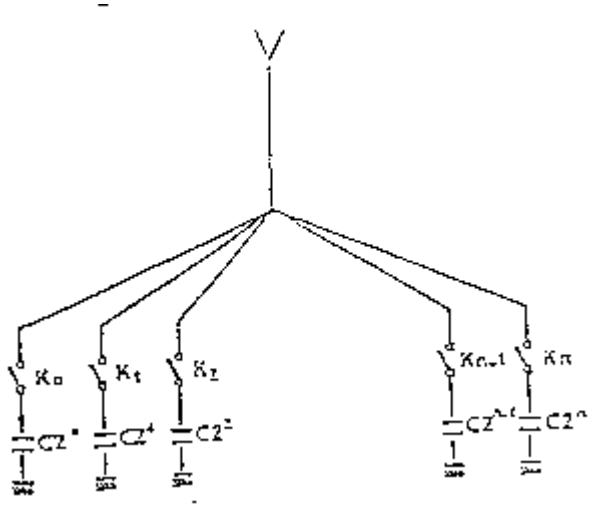


Figure 12

Principle of the capacitor switching

The capacitors which are necessary to tune the antenna reactance are scaled from 3300pF to 60 pF at 2 MHz and 12 MHz respectfully, with a 1,5 pF accuracy at the highest frequencies.

A logarithmic series of n switchable capacitors in parallel defined by Ci=2 Ci-1 with C1=1.5 pF give all discrete value multiple of 1,5pF:

C = S ki Ci from i = 1 to n , with ki= 0 ou 1

C1, which is the smallest used capacitor, defines the accuracy of the C capacitor

The highest individual capacitor value is in theory 3300/2=1650 pF in order to get 3300pF by the addition of all capacitors, and n must be higher than 10.

The total number of capacitors is choosen equal to 12 to takto into consideration the dispersion of the components whose values are guaranteed with a ñ 5% precision, and to recover the possible missing frequency bands.

A special software was created to define and memorize the kiCi arrangements which are necessary to get all discrete capacitor values and recover the possible missing frequency bands. It memorizes the calculated values and the measured values. A calibration at the first installation or in operation in case of a major environment change can be done in less than 6 seconds.

Measured results

The prototype of the antenna achieved a VSWR 2.5:1.

Typical figures are given **FIG13**.

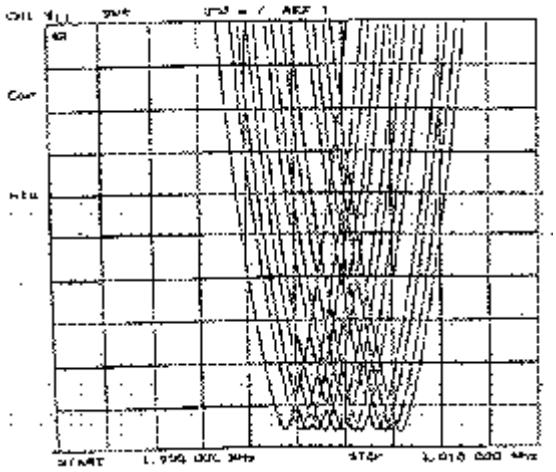


Figure 13

II.7. Qualification

A 2-12 MHz demonstrator was built with acceptable dimensions for land mobile applications (height=90cm, length=2.4m, width=30cm). All adjacent bandwidths were covered within the 2.5:1 VSWR specification.

Its efficiency was measured every 0.5 MHz on a test station by substitution of a refence whip. These values were not more different than ñ1dB from the values deducted from the Q-factor measurements.

A second version with a 2-30MHz frequency range was developped. It was qualified for military environment with mechanical tests (chocks, vibrations) and climatic tests (-40C +70C, rainfall, salted fog, windspeed, ice, dusts, etc...) according to MIL SPEC standards. It is now in service in quantities in the French Army.

III FIELD TRIALS

Extract from the field trial made by Thomson-CSF in octobre 1994 for the French Army:

"From 0 to 600 km, all Q/S and S+N/N measurements have confirmed a behaviour without fault of the half-loop. It always gave results much higher than that of the guyed 5m whip whatever the climatic conditions were (rain, intensive fog...). We tried to use the station in the most extreme environment conditions noting the link results, while driving under the rain, under the high voltage cables either parallel or perpendicular to the road, measuring signal/noise in highly industrialized towns (like Clermont Ferrand) , on the country roads through humid forests, etc...

"The results were independant of these environmental conditions, the reception signal/noise beeing only slightly affected under the very high voltage cables.

"...The half-loop antenna bring the best results in terms of link budget and listening comfort".

Thomson-CSF also confirmed that the half-loop antenna on a moving car allows fast data transmissions without fault in the silent zone of the whip antenna, and that it improves the probability of successfull synchronisation of the new procedures in bad ionospheric conditions.

Other field trials were successfully conducted in France and several foreign countries in the Middle East and America.

IV RADIO INTERFACES

Mobile and naval half-loop antennas and fixed/semi-fixed loop antennas using the same electronic components and softwares are working today with various radiosets for military and civilian applications as well, in frequency hopping, ALE or fixed frequency modes A modular and universal interfacing unit makes it possible to fit the antenna at the radioset RF output using the control interfacing designed for its antenna coupler.

The control exchanges can be done in RS232 or multiwire cable according to the speed. The frequencies can be provided in clear, as a channel number or not provided at all. A frequency counter is necessary in this last configuration.

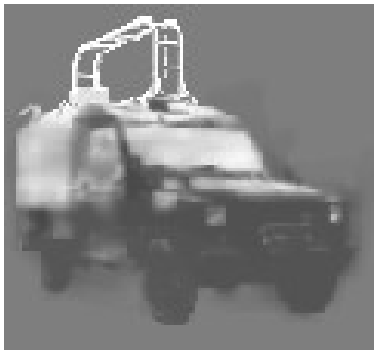
(Continue on the next page)

MILITARY TACTICAL ANTENNAS

STAREC has been involved for a long time in the design of specialized antennas, a wide range.of which has been proved in operation with French and foreign Armed Forces.

This equipment is mainly used in fixed or mobile weapons or telecommunication systems, such as shelters, trucks, battle tanks, forward armoured vehicles, etc. STAREC is involved in the RITA, ROLAND, PR4 G, HF Carthage programs.

Agile half loop on vehicle



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V APPLICATIONS

The chart below present the "not so wellknown" specificities and applications of the HF tuned HF/125W loops and half-loops.

GENERAL CHARACTERISTICS	APPLICATIONS
Very small dimensions for HF (1.5 to 3m rectangle or diameter) Can be radomed	Difficult installations (on roof, small areas, ship,...) Half-loop capability to communicate from a moving vehicle. Discrete stations (fixed and mobile)
Small surface on ground.A ground plane is not necessary for loop	Easy and low cost installation.
Low take-off angle propagation andGround wave radiation (8-shaped pattern)+Near Vertical Incidence Skywave (NVIS) Directivity: + 2dB in free space and +5dB or +6 dB on a conductive ground Gain: - 12/-15 dBi at lowest frequencies to+ 2/+ 5 dBi at highest frequencies.	Communications up to 1000km with 125W, without silent zone in azimuth nor petal nulls in elevation. Communications of the ships along the coasts and over mountains. One antenna only gives the equivalent services of a NVIS antenna (like horizontal dipole) and a vertical whip (at longer ranges). Achievement of an ALE fully automated mobile station: no more need to change antennas at halt alongside the classic silent zone of whips.
High selectivity in the lowest range. High reduction of outband transmissions. High reduction of received noise and improvement of the Signal/Noise ratio(typically 6 to 10dB in reception compared to the wider band antennas like tuned whips or dipoles) High rejection of the strong wideband signals like high voltage lines spurious, indirect effects of lightening, etc... 2 tuned loops are highly isolated (particularly when they are perpendicular with one frame in the central axis of the other one).	Compared to the whip antennas: Better listening comfort, reduced Bit Error Rate (BER) of data transmissions or FH synchronisation signals. Extra filters can be avoided in many applications. Operational in industrial zones and areas of frequent lightning. Simultaneous transmission and reception on the same narrow site (head of a star chained network, duplex station, HF-HF relay,...)
Fully capacitive tuning unit, without coil nor magnetic signature effect	Interesting for certain ships

HF/125W fast tuned frame antennas can find a number of applications for point to point, ground to air and ship to shore applications at any distance to 1000 km.

Old Military HF - Antennas of Communication Cars

Igor Grigorov, RK3ZK
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I have a small collection of information about old military HF antennas used over the World. Presently, three old military automobile HF antennas are described at the article. The antennas are written "as it is," i.e., I give all information, that I have had. I know, the information is not complete at all, but, nevertheless, the information is interesting and it can help somebody to make own 'car antennas.'

Tuned dipole

Tuned dipole exhibited in **Figure 1** was used in army of the USSR. Scheme for the antenna is shown in **Figure 1a**. Tuned dipole made from a strong tube that has diameter 15 to 30 mm. The tube is installed at height of 1-1,5 m above the roof of an automobile and goes out approximately on 1 meter for overall dimensions of the automobile, as it is exhibited in **Figure 1b**. Tuned dipole has high radiation both at low and high (mostly) angles. It allows the antenna to make links by earth and reflected from ionosphere wave.

Figure 1 Tuned dipole

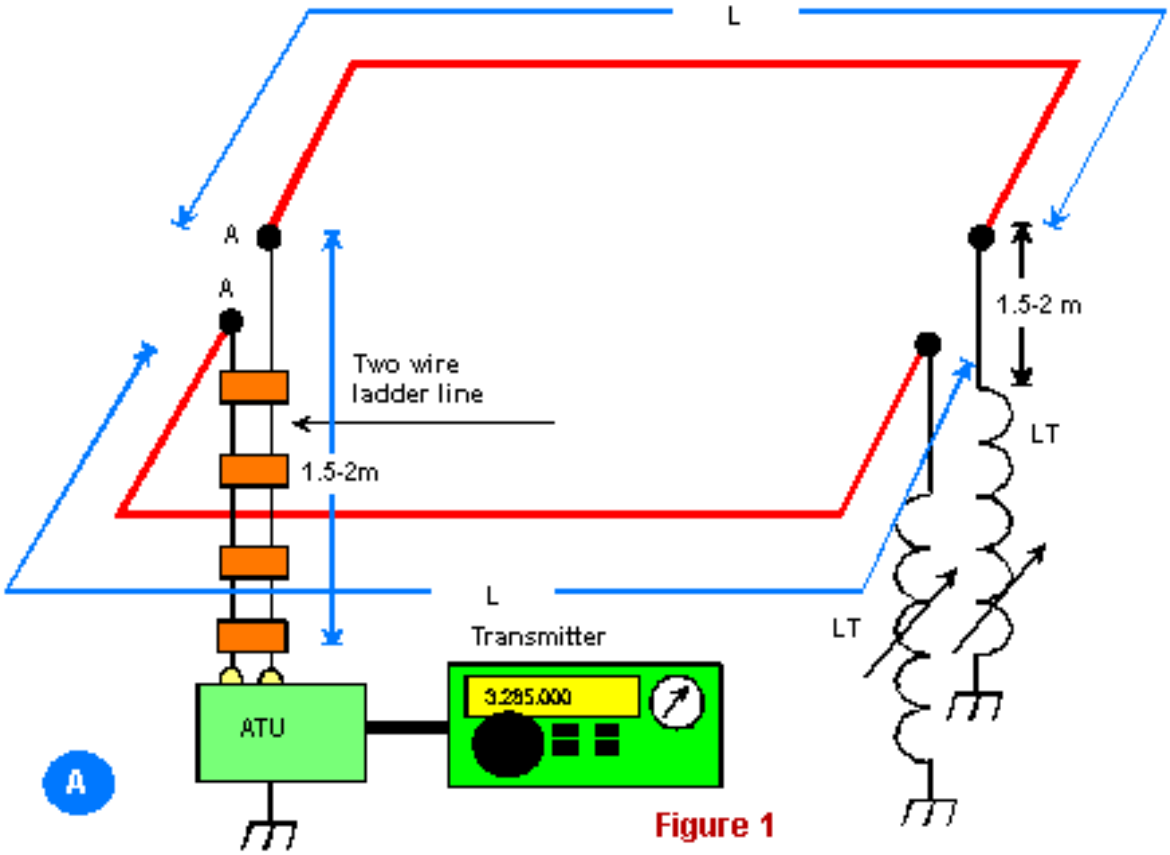


Figure 1

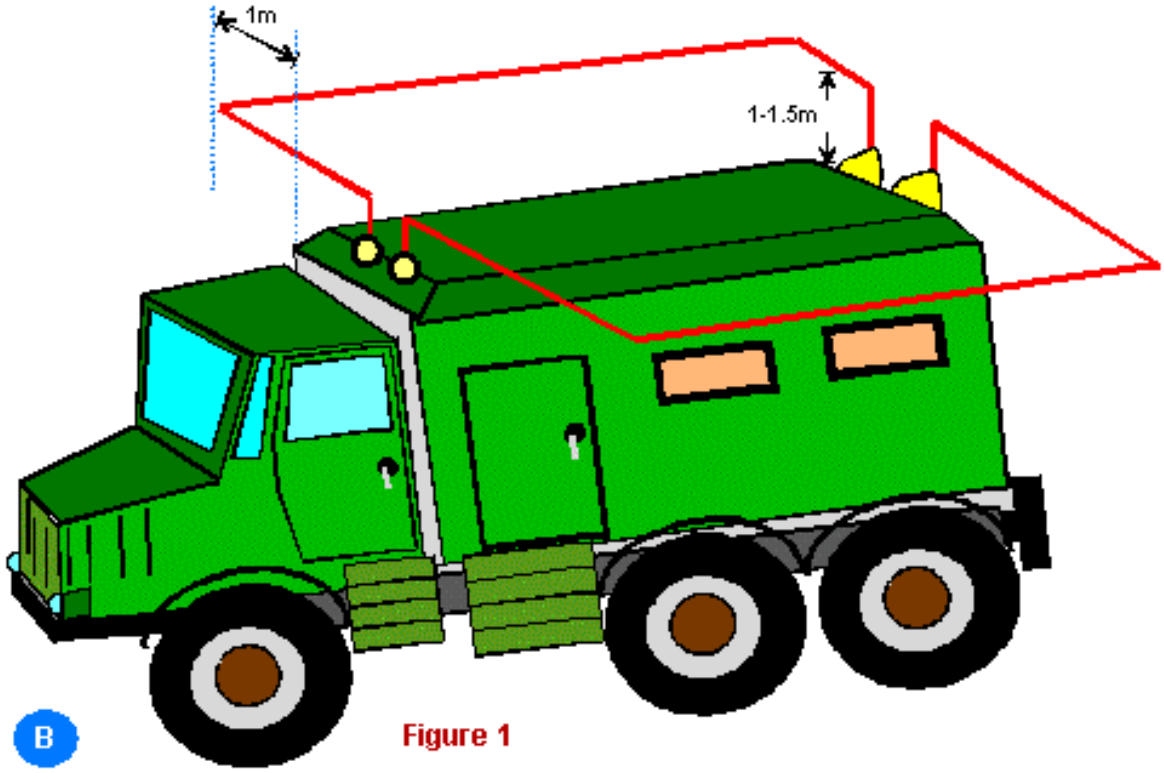


Figure 1

Length 'L' of the antenna (see **Figure 1a**) depends to dimensions of a car, where the antenna is installed, usually the length close to 6 meters. The antenna feeds with help a short length of a two-wire line, usually the length is equal to 2- 3 meters, an ATU is used. Antenna is adjusted with help of a symmetrical variometer 'LT' installed at opposite ends of the tuned dipole. The antenna is tuned on maxima of RF current to points 'A'. In such case the antenna ensures maximum effective work. But in the manual of the antenna is pointed, that the antenna in some cases can be tuned on maxima of RF voltage on points 'A'. Such set-up for the antenna is possible if the communication car is placed on a good conducting surface (it can be moist salty soil) or by operation from natural shelters - holes, ravine. Inductances of antenna variometer vary from several microhenry up to 300 microhenrie. The antenna works good at 2-25 MHz.

At usage of this antenna in military communication car, the antenna gives that advantage, that the roof of the automobile remains free. It enables to install on the roof other antennas, for example, for VHF-UHF ranges

Folded dipole

Folded dipole was in use within the World War – II and till 70s of the 20 century. The dipole is a wire folded by meander and loaded to serial coil plus a capacitor. **Figure 2** shows the disposition of the folded dipole on communication car. The antenna is located at the altitude approximately at 1-1,5 meters above the roof. **Figure 3** shows the scheme of the antenna. Wings of the folded dipole could reach to 10-15 meters in length. It depends on sizes of the car.

Folded dipole is fed by two-wire ladder line in length about 3 meters. The line is connected to an ATU. The antenna is tuned on maxima of RF current to points 'A'. Folded dipole is tuned in resonance in the operation frequencies with help of loading spools 'LT' together with capacitor 'C', that made as a constructing part of the car.

The antenna is intended for 60-90 meters, and usually does not work at other ranges. . Folded dipole has strongly radiation to the sky and a little to the horizon.

Dipole with low characteristic impedance

Dipole with low characteristic impedance is intended for a work at wide frequencies range and for installation on the roof of a communication car that has small dimensions. **Figure 4** shows the scheme (**Figure 4A**) and disposition (**Figure 4B**) of the dipole on communication car. I must say, that I have seen some photos, where the antenna was installed athwart to the roof, as it is shown at **Figure 4C.** The antenna is located at the altitude approximately at 1 meter above the roof. Wings of the dipole made as a metal grid has shape, crosswire at the cells soldered. The wing of the dipole has the width in (0,5-1,5)-meters and the length (1,5-2)-meters.

Figure 5 shows the scheme of feeding of dipole with low characteristic impedance. The antenna is connected through a two-wire ladder line in about 2 meters length to an ATU. The ATU has a resonance

Figure 2 *Folded dipole placed on communication car*

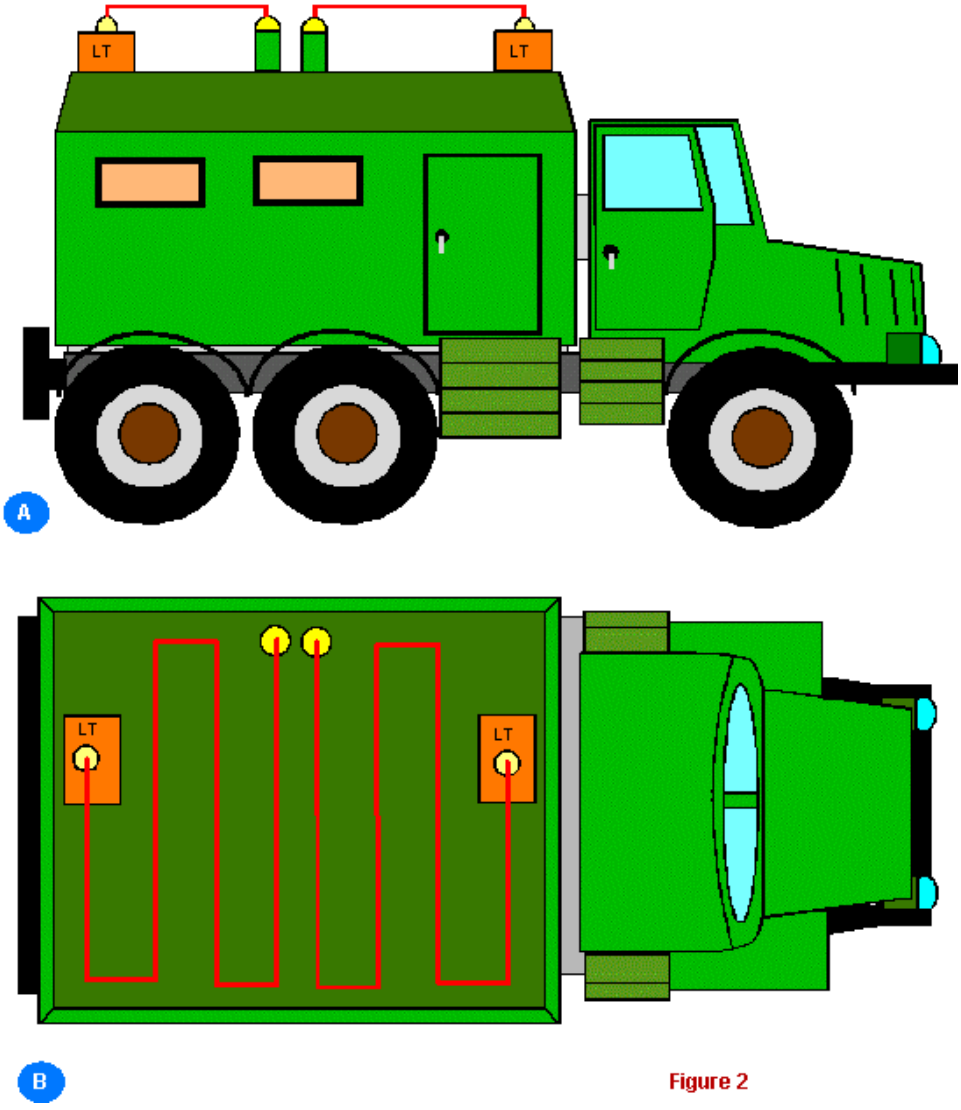


Figure 2

Figure 3 *Scheme of the folded dipole*

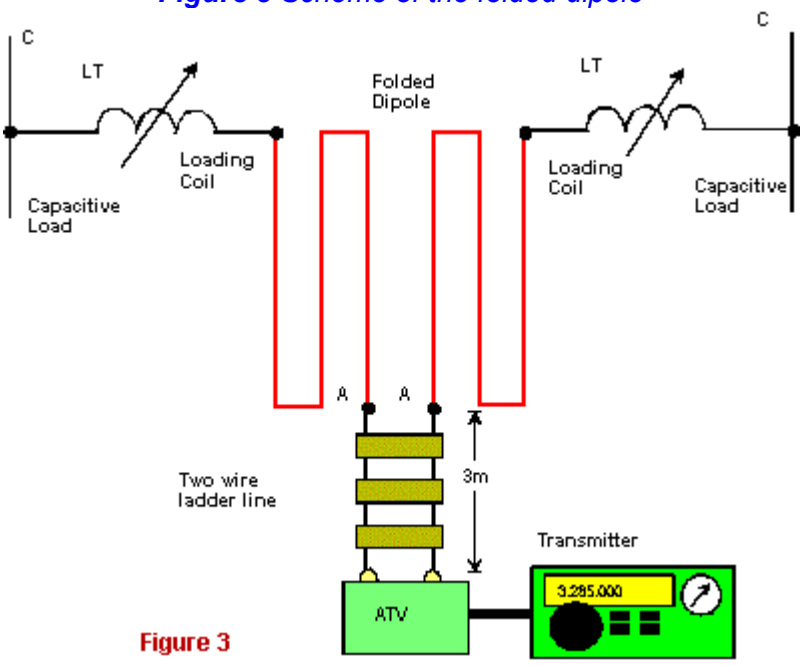


Figure 3

Figure 4 Dipole with low characteristic impedance

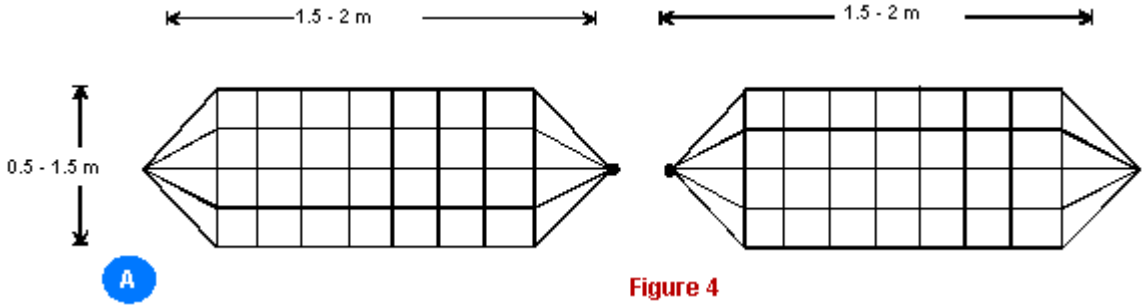


Figure 4

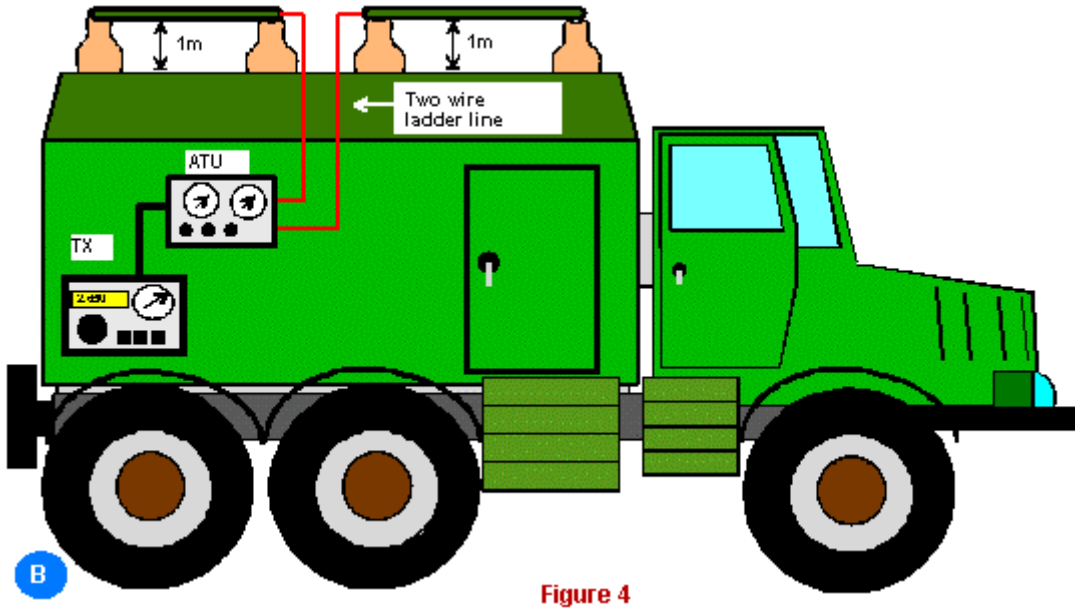


Figure 4

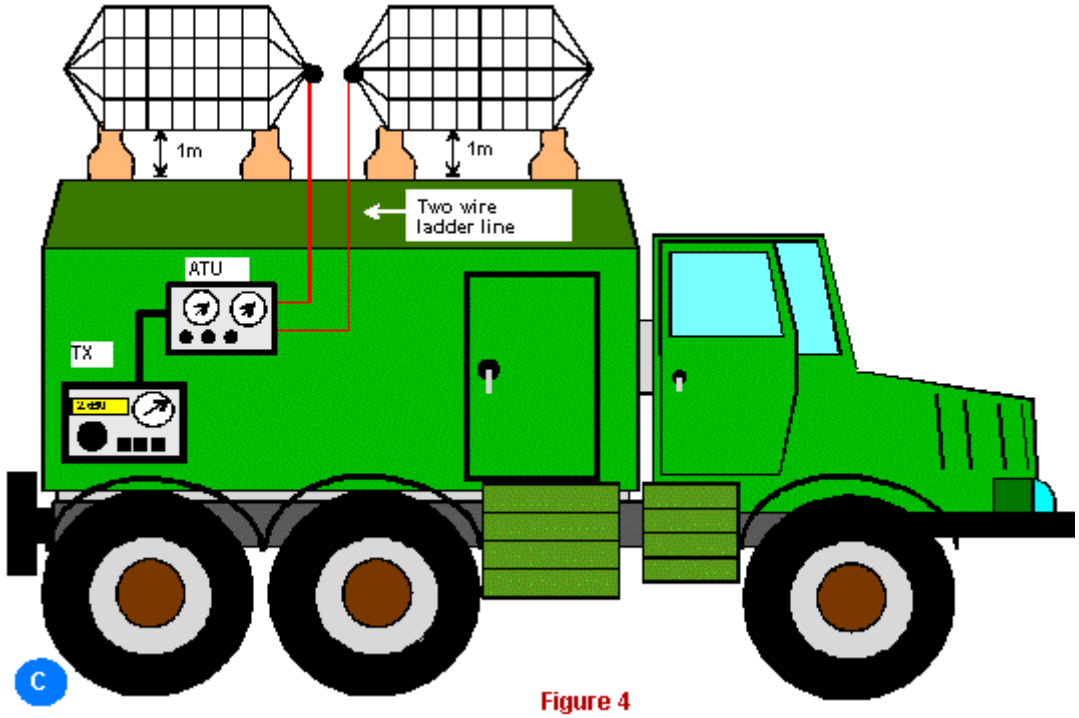


Figure 4

Figure 5 Scheme of feeding of dipole with low characteristic impedance.

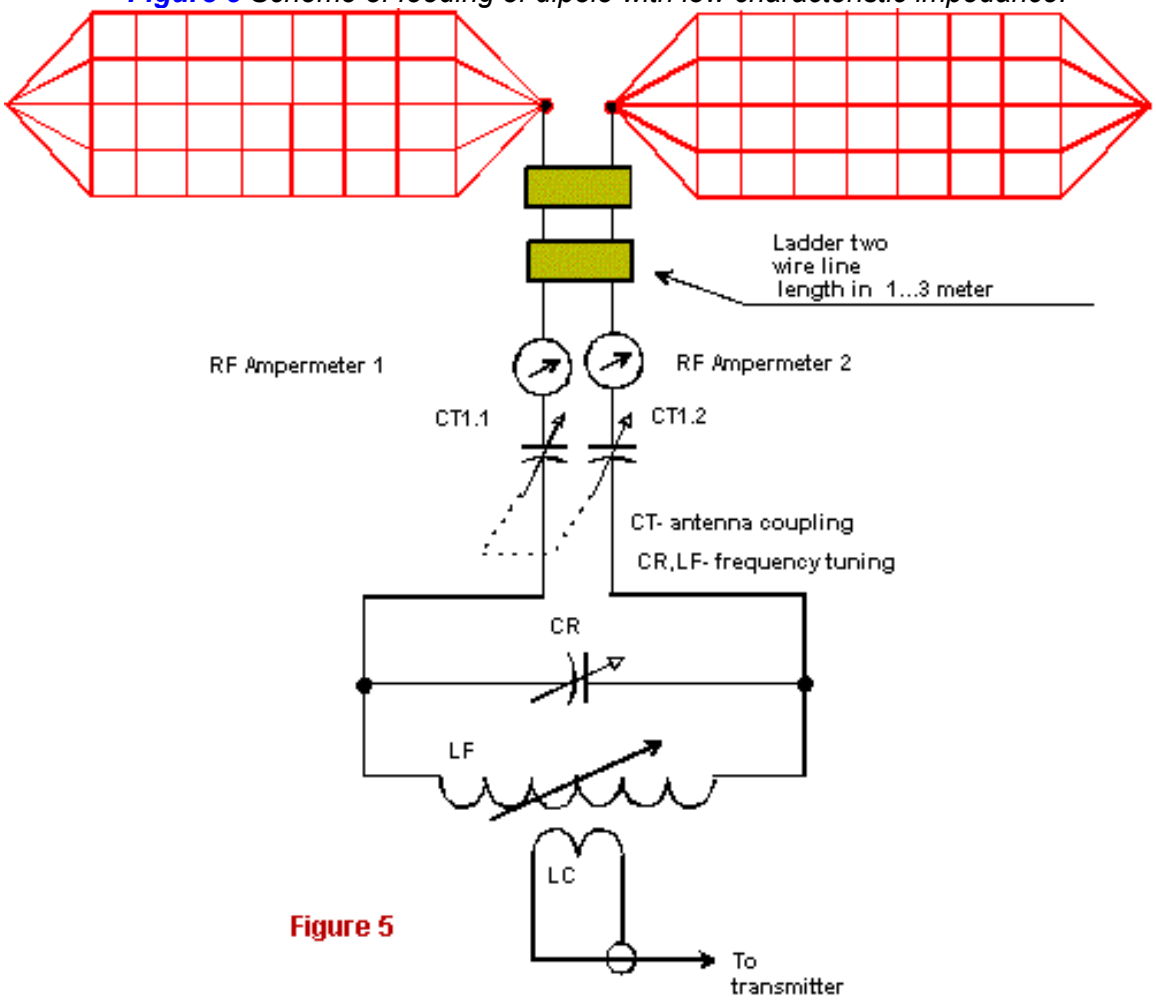


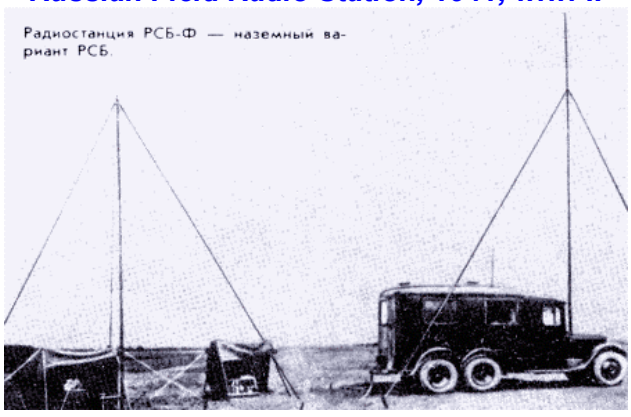
Figure 5

circuit 'LF- CR', that is coupled with help of coupling coil 'LC' with a transmitter. With the help of variable
spool 'LF' and variable capacitor 'CR' this circuit can be retune in the frequencies range of 2-20 MHz. In the same frequency range the antenna works. Maxima of RF current in to the antenna is installed with the help of variable symmetrical capacitor 'CT.'

Such antenna was widely used in communication cars during the World War-II and some time after the war. The antenna ensures the sure communication with zenith radiation in HF range 2- 4 MHz in radius of 200-300 km from the antenna. Also this antenna ensures long-distance communication in HF ranges 5-20 MHz.

But this antenna had the deficiencies: at the first, it has too complicated matching device, at the second the antenna takes too much place in the roof of the car. For these reasons, since of the end of 50s, the antenna practically is not used in military communication cars.

Russian Field Radio Station, 1941, w.w.-II



Russian Field Radio Station, 1913, w.w.-I

Modern Military HF - Antennas of Communication Cars

Igor Grigorov, RK3ZK
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The basic types of military *antennas, which for a long time were used on communication cars, were construed during and before the World War II. After WW-II researches for new antennas for communication cars was renewed. Below we shall consider two new type of car antennas which have appeared in army after the World War-II still are in use on modern communication car. There are magnet antennas and DDRR antennas, that began to be used for communication cars (under my information) rather recently - in the middle of 70s years of the 20 century.*

Mag Loop antennas of communication cars *for 150-80 meters*

As usual a magnet antenna of the range is installed on a communication car as it is shown in **Figure 1**. The magnet antenna ensures sure communication in radius of 200 kms at the daylight time and up to 400 kilometers at the night time.

Figure 1 Magnet antenna on a communication car

The magnet antenna at marching condition is installed parallel to the car roof (see. **Figure 1a**) and does not hinder to ride the car under low bridges or under trees in forests. The magnet antenna stands in operating position with the help of an electric motor or by operator hand. **Figure 1b** shows the zenith magnet antenna in the operating position.

The magnet antenna (item 2, **Figure 1b**) is tuned in resonance on the operating frequency with the help of variable capacitor (item 3, **Figure 1b**), which is weatherproofed by hermetic box (item 4, **Figure 1b**). The variable capacitor is turned with the help of an electric motor. The magnet antenna is drove by a small loop (item 1, **Figure 1b**) that is installed in corner of the magnet antenna. Driven loop is coupled to the transmitter with the help of a coaxial cable by characteristic impedance of100 Ohm. The sizes of the army magnet antenna for operation in 150 - 80 meters are exhibited in **Figure 1b**.

Mag Loop antennas of *communication cars* for meters

Also magnet antennas are used for operation in HF ra 60 meters. Such magnet antenna has smaller contrasted to magnet antenna intended for operation meters. Magnet antenna for 90-60 meters owing to sizes is installed or above the roof of the cabin of comn

car (see **Figure 2**) or at back edge of the car (see **Figure 3**). At this installation of mag loop the roof of the car is free to place other antennas or some more electronic equipment. The sizes of the magnet antenna for operation in 90 - 60 meters are exhibited in **Figure 2** and **Figure 3.**

Figure 2 Magnet antenna above the roof of the car cabin

Figure 3 Magnet antenna on the back edge of the car

The magnet antennas shown in **Figures 1-3** usually are made of an aluminum bent tube in 20- 40 millimeters in OD and 2-3 millimeters thick.

Magnet antennas for 150 -90 meters is not intended for operation when a communication car is moving. But Magnet antennas for 90 - 60 meters can be used when a communication car is moving.

Magnet antenna on a communication car

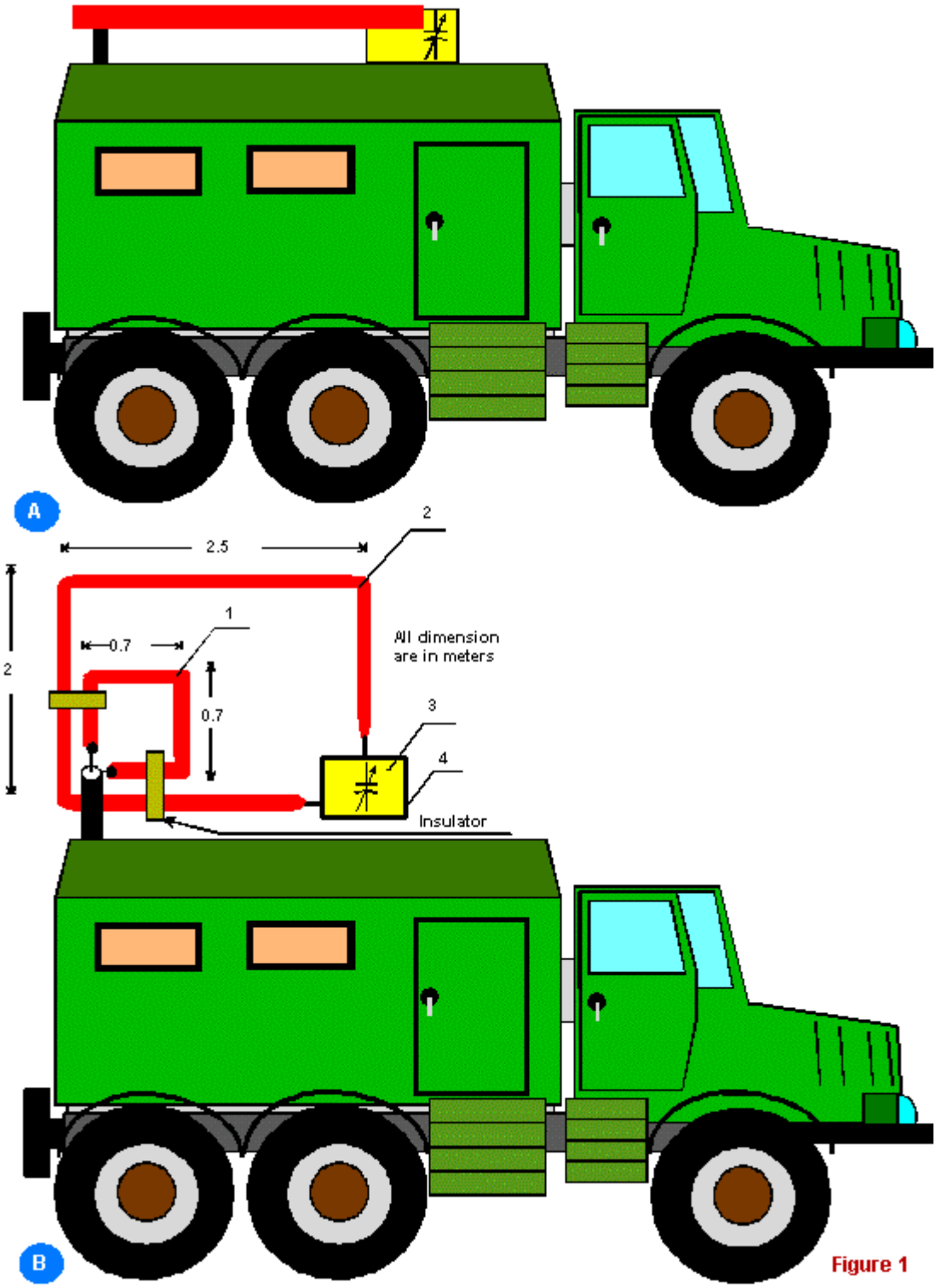


Figure 1

Antennas DDRR in military communication

Antennas DDRR are used in military communication of many countries. **Figure 4** shows the schematic of antenna DDRR. Say simple the antenna represents a tube bent in shape of letter "L". The length of the 'L'

Is little smaller the quarter wave of the high lower range of the antenna, but the length 'L' not less the lower operation range of the antenna. Antenna DDRI to resonance by the variable capacitor C which is dri electric motor M. Antenna DDRR, as usual, is mad aluminum tube in 20-40 millimeters OD.

Figure 4 Schematic of Antenna DDRR of communicati

Magnet antenna above the roof of the car cabin

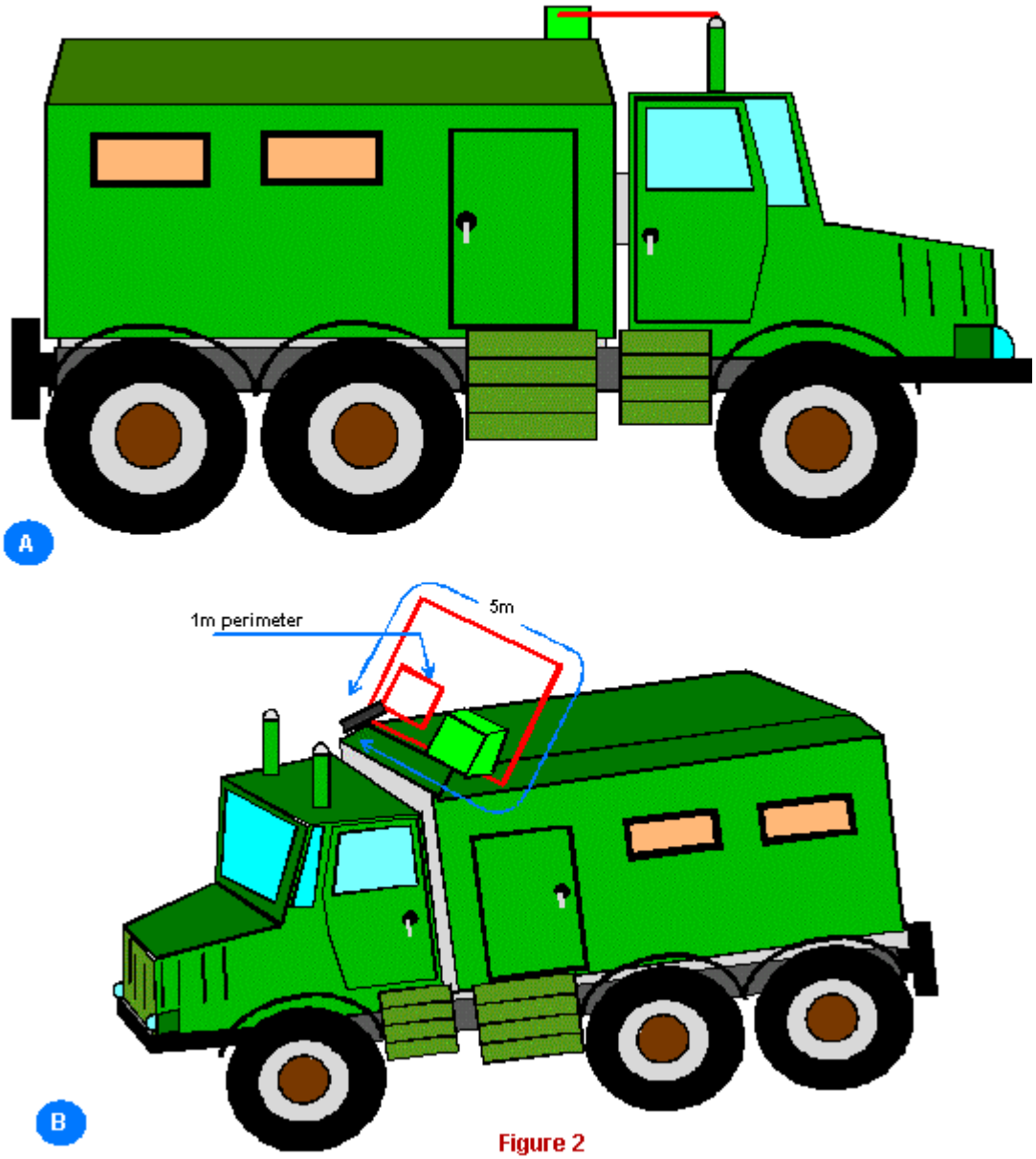


Figure 2

On military vehicles an antenna DDRR usually places on one of edges as it is shown in **Figure 5**. It allows to install other antennas on the roof of the communication car. Antennas DDRR often serve as enclosure of the roof.

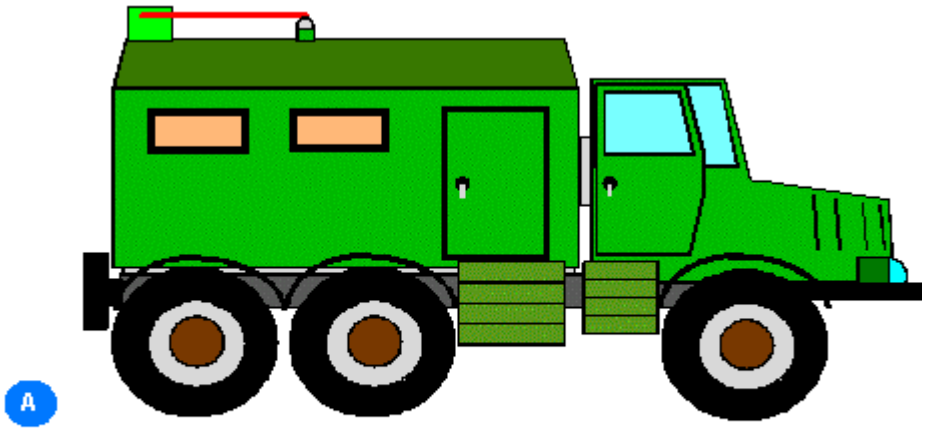
Figure 5 Antenna DDRR on military vehicle

Antennas DDRR also are used on some heavy tanks or BMI (battle machine of infantry). On this military vehicle the antenna DDRR serves as a rail for other hand. Antenna DDRR has mechanically strong design, so it is difficult to damage an antenna DDRR in battle.

Seldom use two antennas DDRR that fed with some phase shift to create special diagram directivity. **Figure 6** shows such antenna array.

Figure 6 Array of antennas DDRR

Antenna DDRR is very strong and allows to do comn when the car is moving. Radio amateurs also can us DDRR for the operation in ether at installation on the rc car.



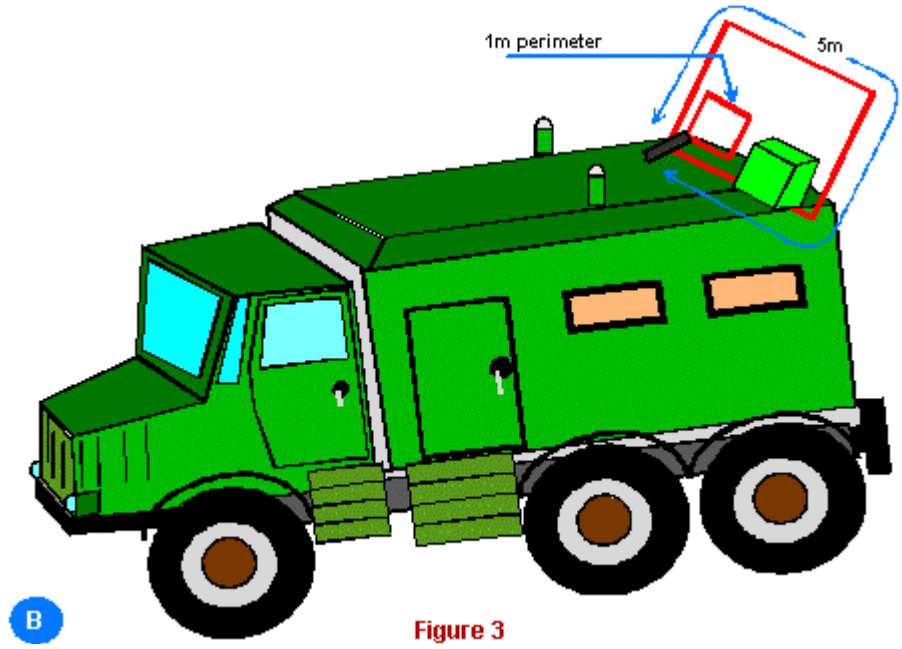


Figure 3

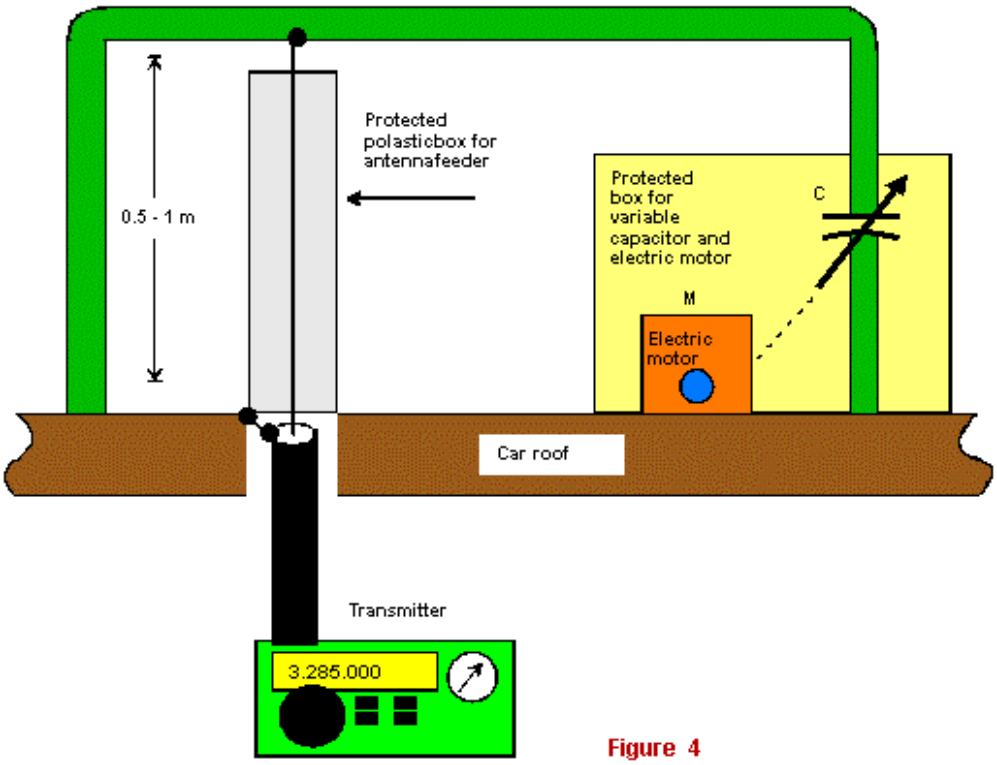


Figure 4

Antenna DDRR on military vehicle

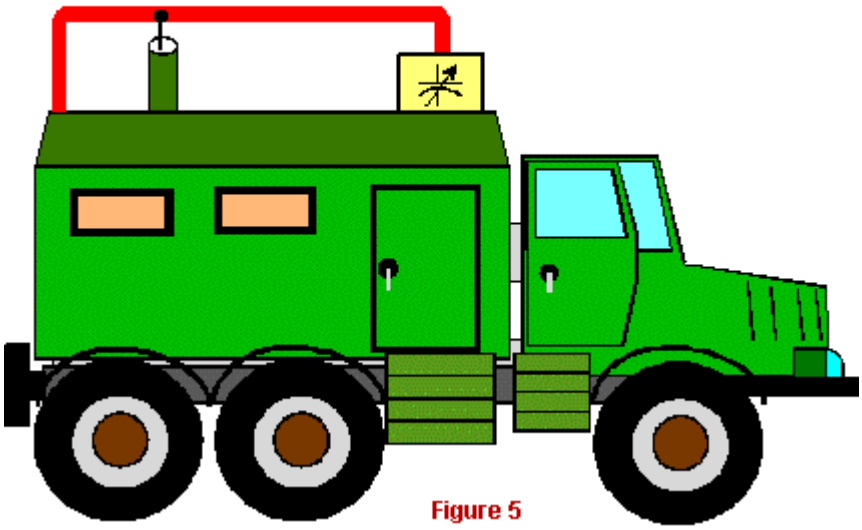


Figure 5

Array of antennas DDRR

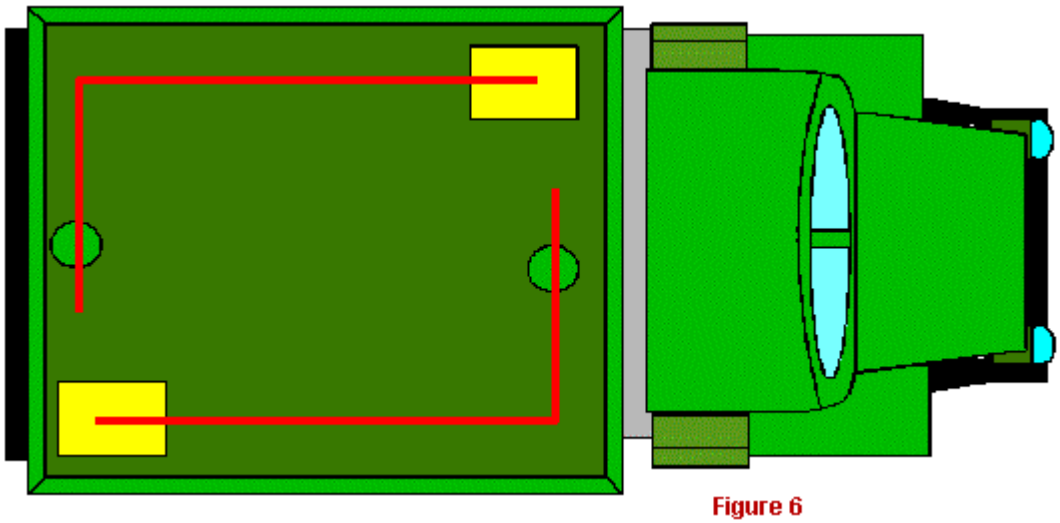


Figure 6

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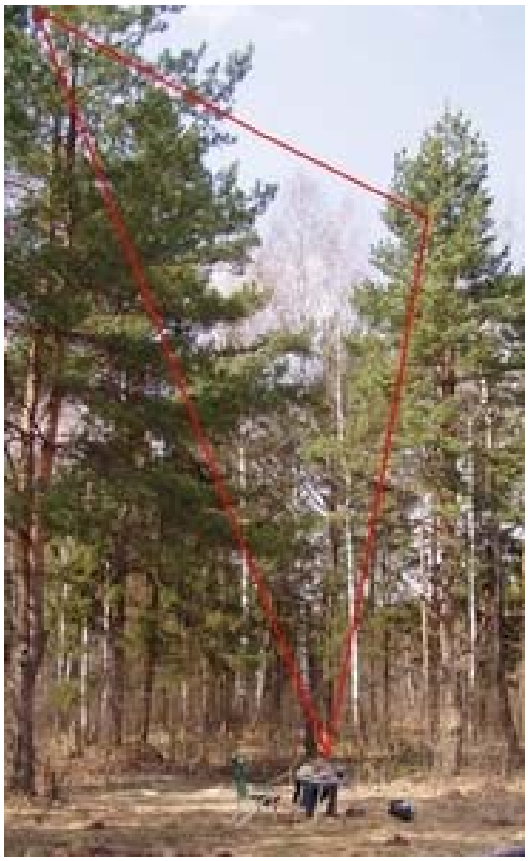
In Russia summer traditionally is the season for radio expeditions. So, the question about a field antenna stands before hundreds radio amateurs. Certainly, it is impossible to give one answer to this question. A design of an antenna for fields depends on many factors. There are frequencies bands used by radio expedition, local factors for antenna installation, time and money and so on. Igor, RV3DA, has developed universal wire antenna. This one with ATU works well on all amateurs short-wave ranges, including WARC. factors. There are frequencies bands used by radio expedition, local factors for antenna installation, time and money and so on. Igor, RV3DA, has developed universal wire antenna. This one with ATU works well on all amateurs short-wave ranges, including WARC.



The antenna has a triangular shape, one of the ends of the triangular is grounded. **Figure 1** shows the circuit of the antenna. A good grounding is necessary for successful work of the antenna. However, the antenna provides good work without good grounding also.

The antenna was tested by team of collective radio station RK3DZD in field conditions.

On Eastern Sunday April 11 we have been going to RDA- pedition. About RDA program you can see ***page 86, ANTENTOP# 2- 2004***. Our team (RD3DT, UA3DUS, RZ3DT and RV3DA) and buys settled down at picturesque surroundings of the edge of Tsna river, Egor'evskoe area **MO-62** at Moscow oblast.

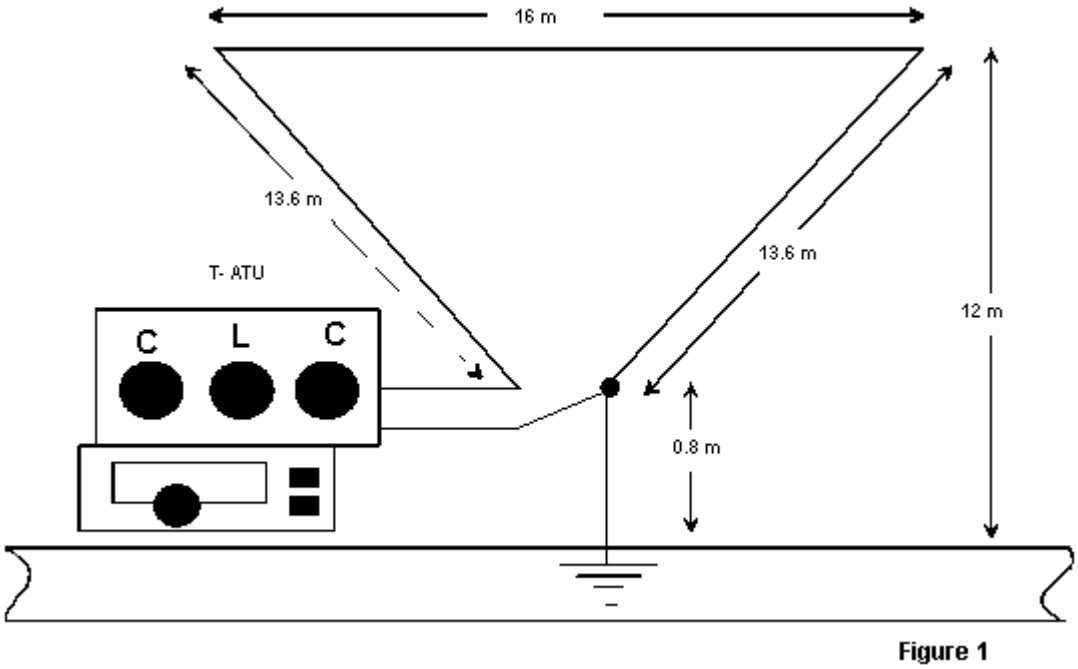


Figure 1

Yuri, RA3DUF, hanged our antenna on the nearest pines. TNX to Yuri, RK3DUF for steeplejack works! You can see a photo of the antenna (antenna colored by red). The antenna had no feeder and was connected directly to a transceiver FT-847 through a home- made T- ATU that was placed at the table. We used an automobile accumulator 55-A/h and gas-generator Honda for feeding our equipment.

Grounding was very bad - a pipe in diameter of ½" and in length of 1 meter. The ground was very damp, ten meters father from the antenna a river was. The antenna was hanged between two high pines.

What we have had:

1. 40 meters - is higher than any praises. Really, we ruled by pile- up from tens calling stations. Simultaneously we received as local as DX- stations.

2. 20 meters – the directivity of the antenna did bad effect for us. The antenna was directed to the North - South, that it was not good for us. Southern radio stations simply rattled in the Air!

3. 15 meters- Japanese called us very loudly.

4. On 10 meters was dead (no propagation).

5. 80 meters – we received very loudly many radio station from Siberia (2000- 3000 kms from us) but our sign was received poor in Siberia because, as we think, of poor grounding.

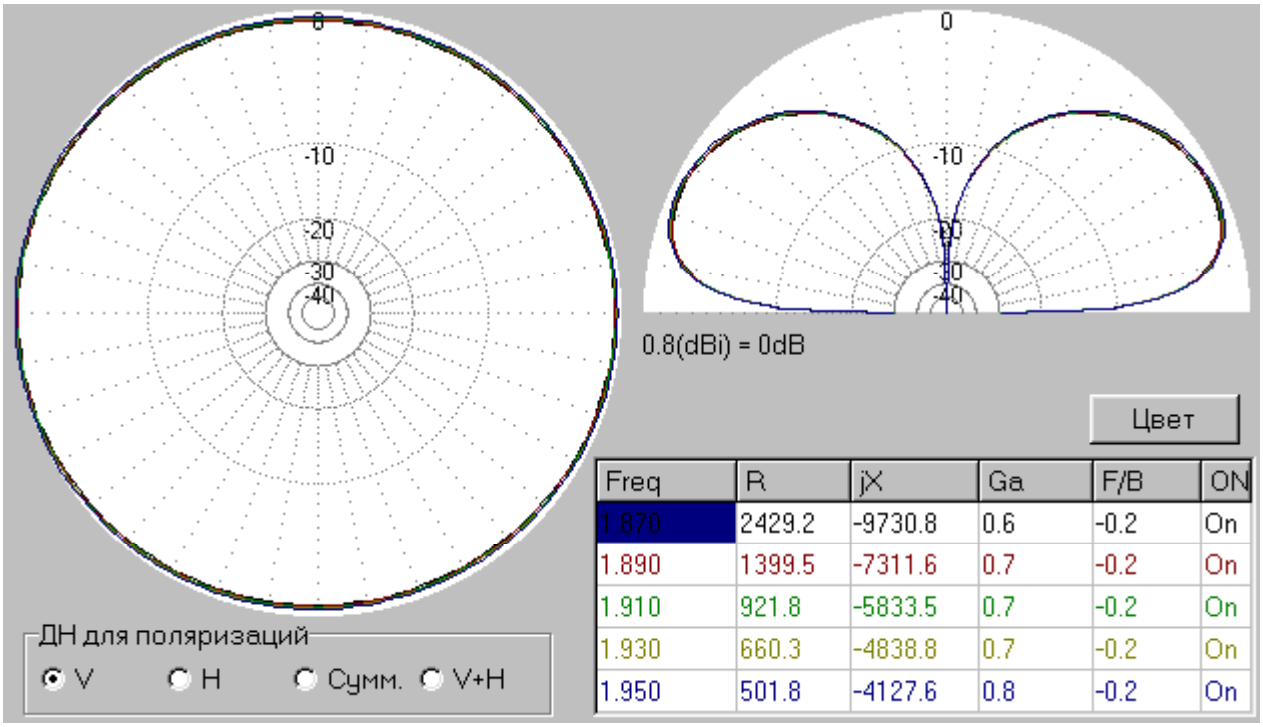
Below given diagram directivity for the antenna obtained with help of free antenna program MMANA (MININEC based). Left diagram is a section of the volumetric diagram directivity of plane X-Y at a zenith corner of the maximum radiation. The right diagram is section of the volumetric diagram directivity of plane X-Z. Also at the right down corner of the pictures is a table with antenna impedance. Please, take attention to the data, you can do decision how you ATU does match of the

73!
Igor Grigor'ev, RV3DA

So our first workplace looked. As usual, we had too little time for preparing before our pedition, so, we have used "table" T-ATU. Two air variable capacitors and a variable inductor placed on the table without any case. Yury, RK3DUF, did DX- QSO.

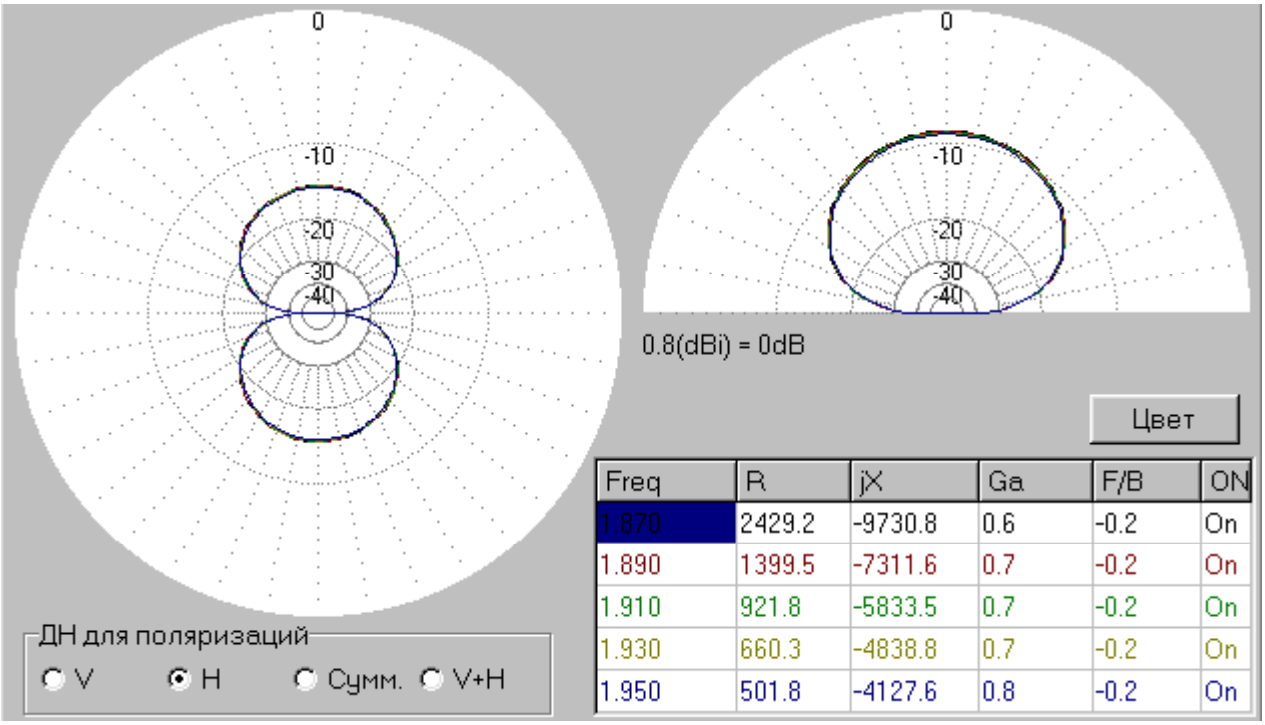
Antenna RV3DA at 160-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
1.870	2429.2	-9730.8	0.6	-0.2	On
1.890	1399.5	-7311.6	0.7	-0.2	On
1.910	921.8	-5833.5	0.7	-0.2	On
1.930	660.3	-4838.8	0.7	-0.2	On
1.950	501.8	-4127.6	0.8	-0.2	On

Horizon Radiation Pattern

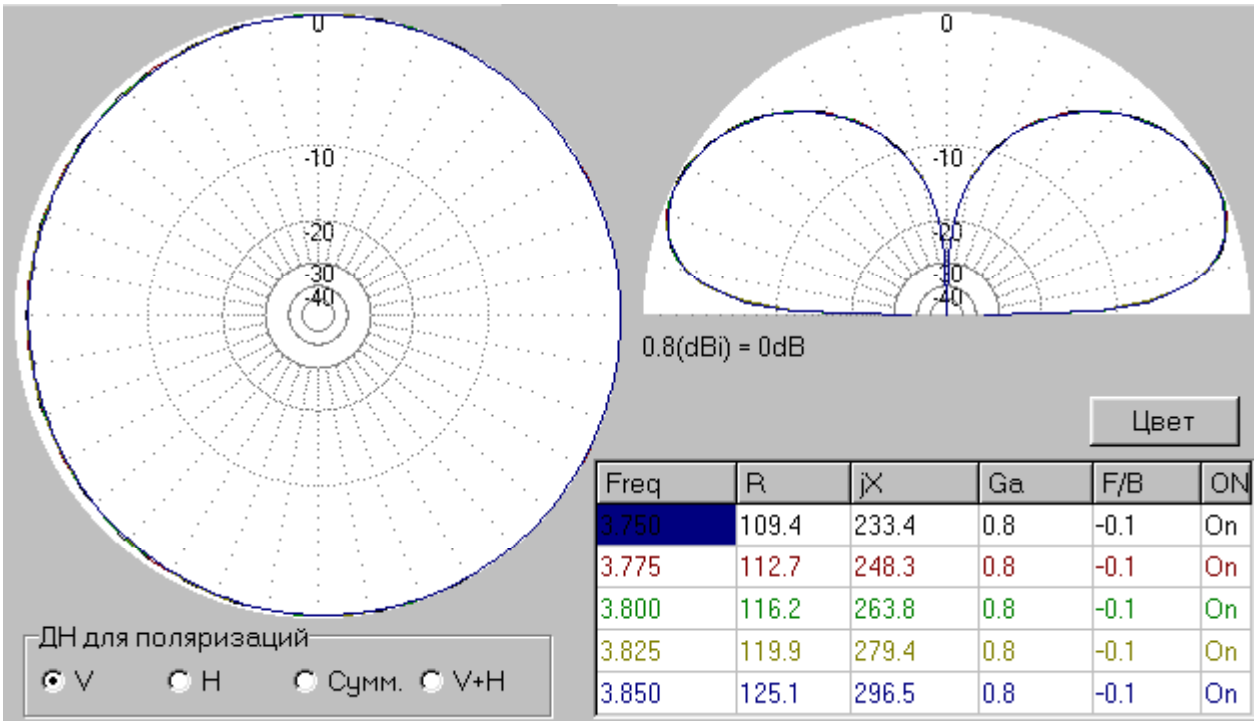


Freq	R	jX	Ga	F/B	ON
1.870	2429.2	-9730.8	0.6	-0.2	On
1.890	1399.5	-7311.6	0.7	-0.2	On
1.910	921.8	-5833.5	0.7	-0.2	On
1.930	660.3	-4838.8	0.7	-0.2	On
1.950	501.8	-4127.6	0.8	-0.2	On

Comments: Antenna radiates mainly radio waves with vertical polarization. A very good pattern with low lobes in the vertical plane. Circular pattern in horizontal plane. It is fine for DX- QSO. But antenna has Z= 921-j5833-Ohms at 1910-kHz. Not all ATUs do good matching for such load.

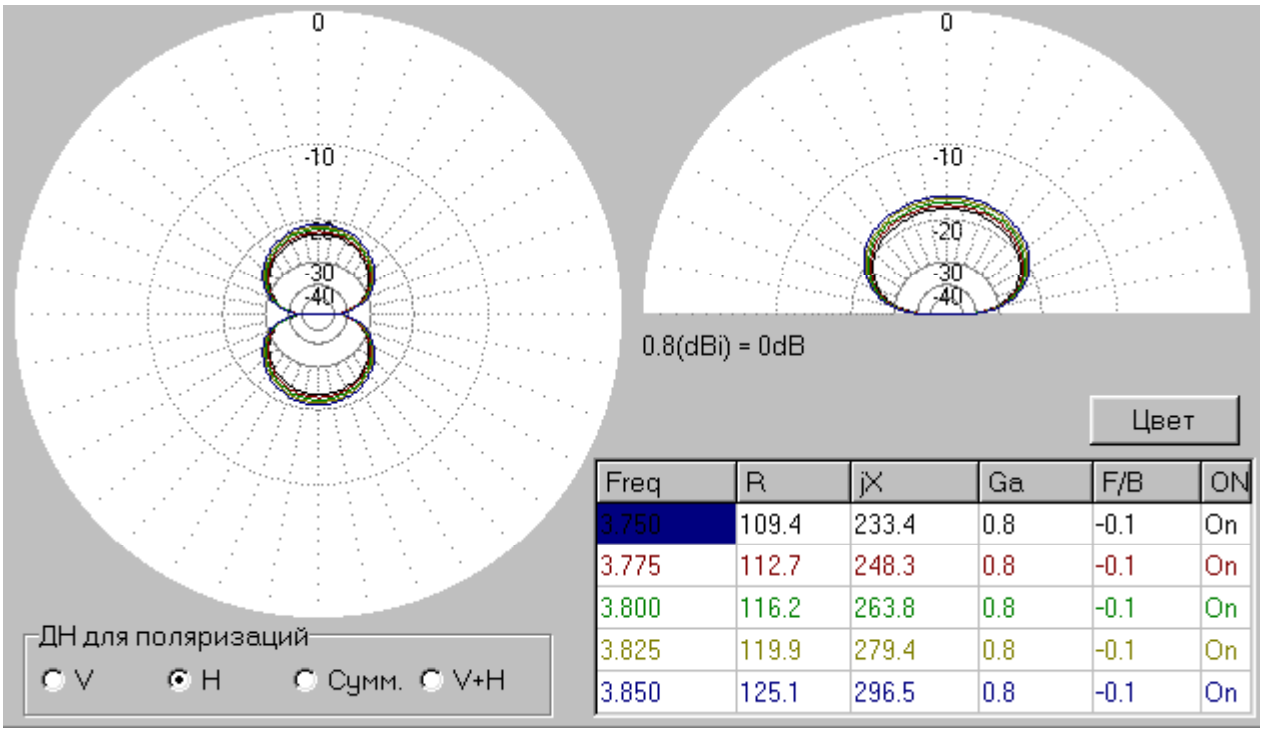
Antenna RV3DA at 80-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
3.750	109.4	233.4	0.8	-0.1	On
3.775	112.7	248.3	0.8	-0.1	On
3.800	116.2	263.8	0.8	-0.1	On
3.825	119.9	279.4	0.8	-0.1	On
3.850	125.1	296.5	0.8	-0.1	On

Horizon Radiation Pattern

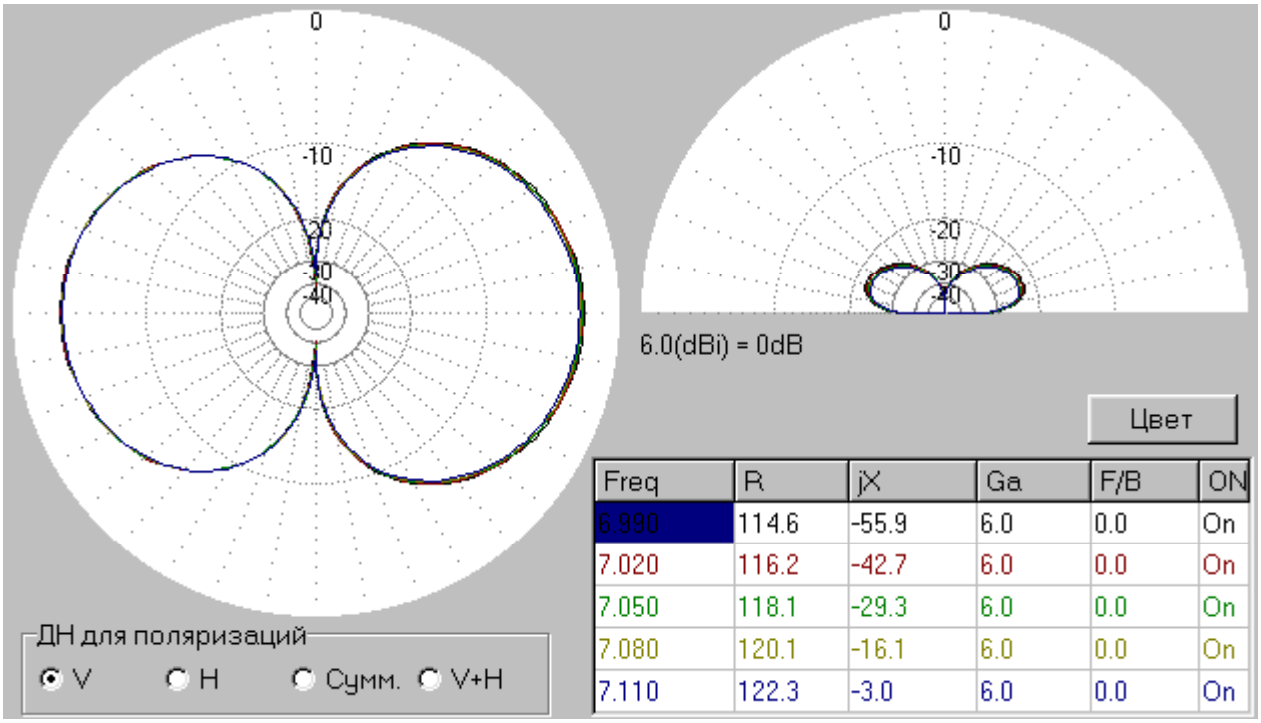


Freq	R	jX	Ga	F/B	ON
3.750	109.4	233.4	0.8	-0.1	On
3.775	112.7	248.3	0.8	-0.1	On
3.800	116.2	263.8	0.8	-0.1	On
3.825	119.9	279.4	0.8	-0.1	On
3.850	125.1	296.5	0.8	-0.1	On

Comments: Antenna radiates mainly radio waves with vertical polarization. A very good pattern with low lobes in the vertical plane. Circular pattern in horizontal plane. It is fine for DX- QSO. Antenna has Z= 116+ j263-Ohms at 3800-kHz. Almost any ATU does good matching for such load.

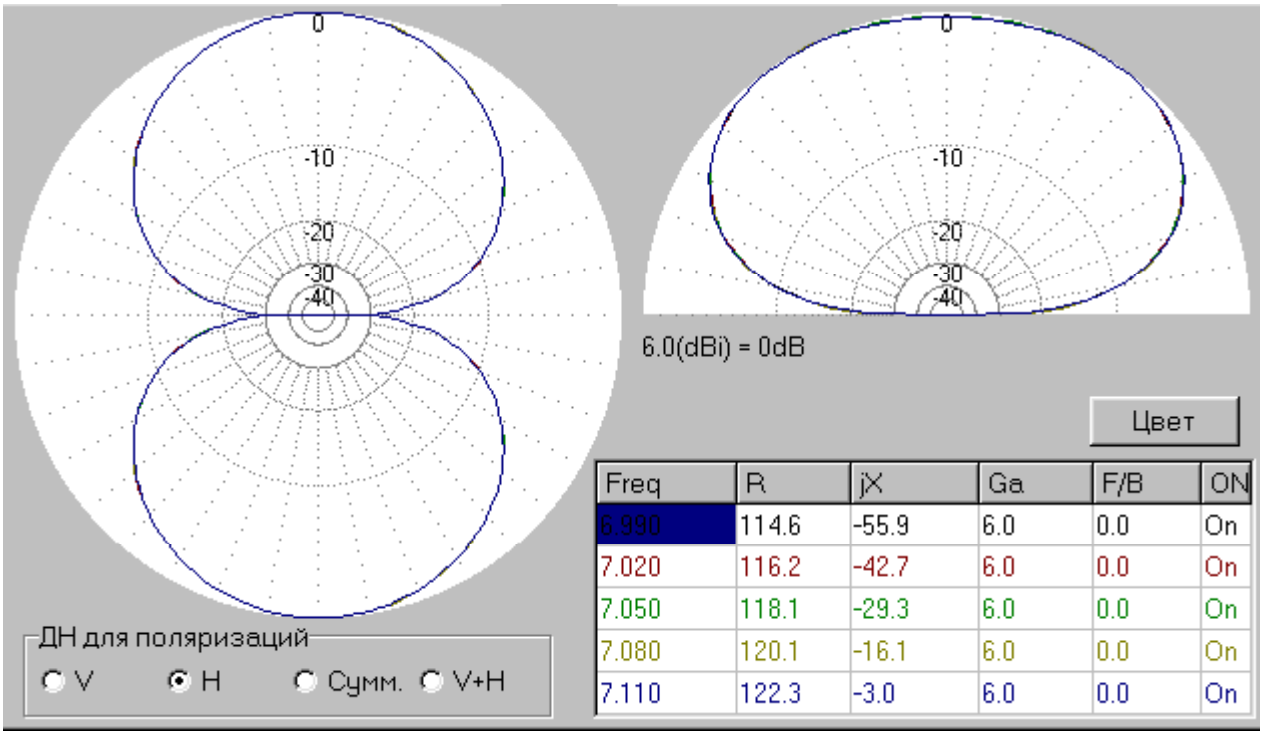
Antenna RV3DA at 40-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
6.990	114.6	-55.9	6.0	0.0	On
7.020	116.2	-42.7	6.0	0.0	On
7.050	118.1	-29.3	6.0	0.0	On
7.080	120.1	-16.1	6.0	0.0	On
7.110	122.3	-3.0	6.0	0.0	On

Horizon Radiation Pattern

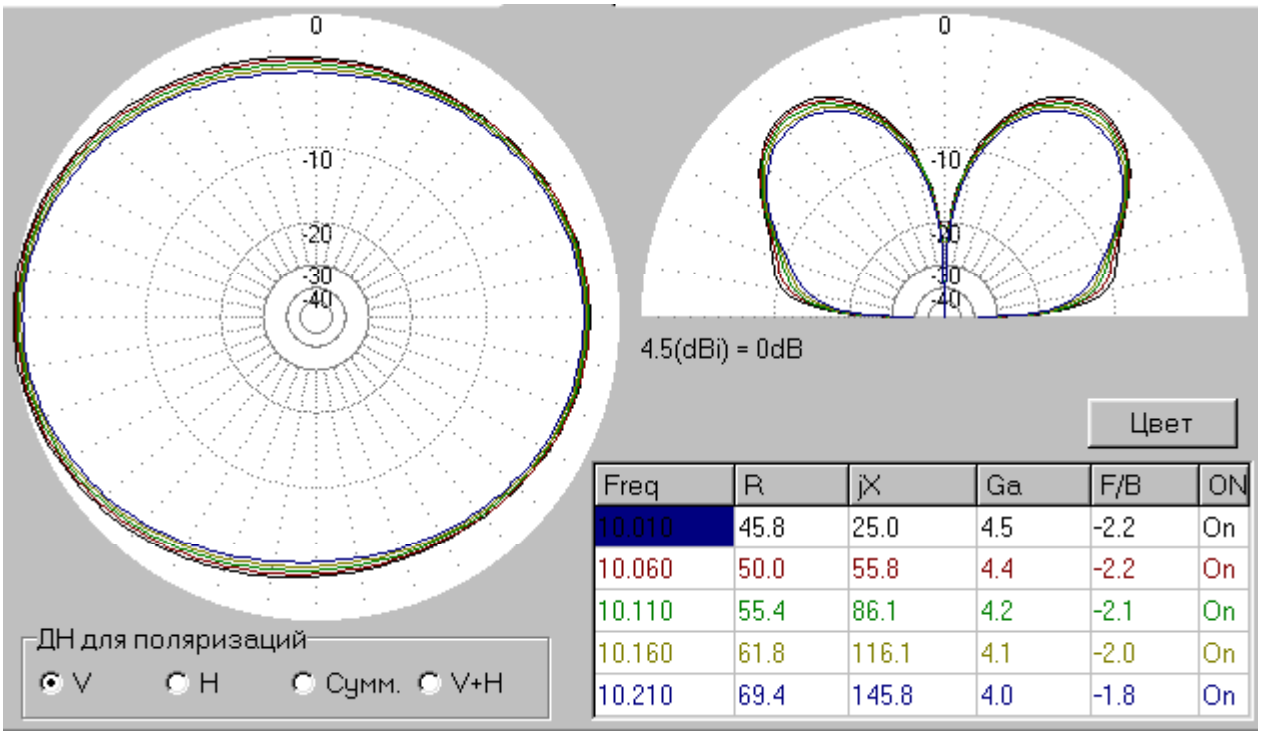


Freq	R	jX	Ga	F/B	ON
6.990	114.6	-55.9	6.0	0.0	On
7.020	116.2	-42.7	6.0	0.0	On
7.050	118.1	-29.3	6.0	0.0	On
7.080	120.1	-16.1	6.0	0.0	On
7.110	122.3	-3.0	6.0	0.0	On

Comments: Antenna radiates mainly radio waves with horizontal polarization. A good pattern in the vertical plane. Antenna has strong zenith radiation that allows to do local QSOs. “Eight- figure” pattern in horizontal plane, so, it demands to choose a proper direction before an installation of the antenna. Antenna is fine for DX and local QSOs. Antenna has Z= 116+ j263-Ohms at 3800-kHz. Almost any ATU does good matching for such load.

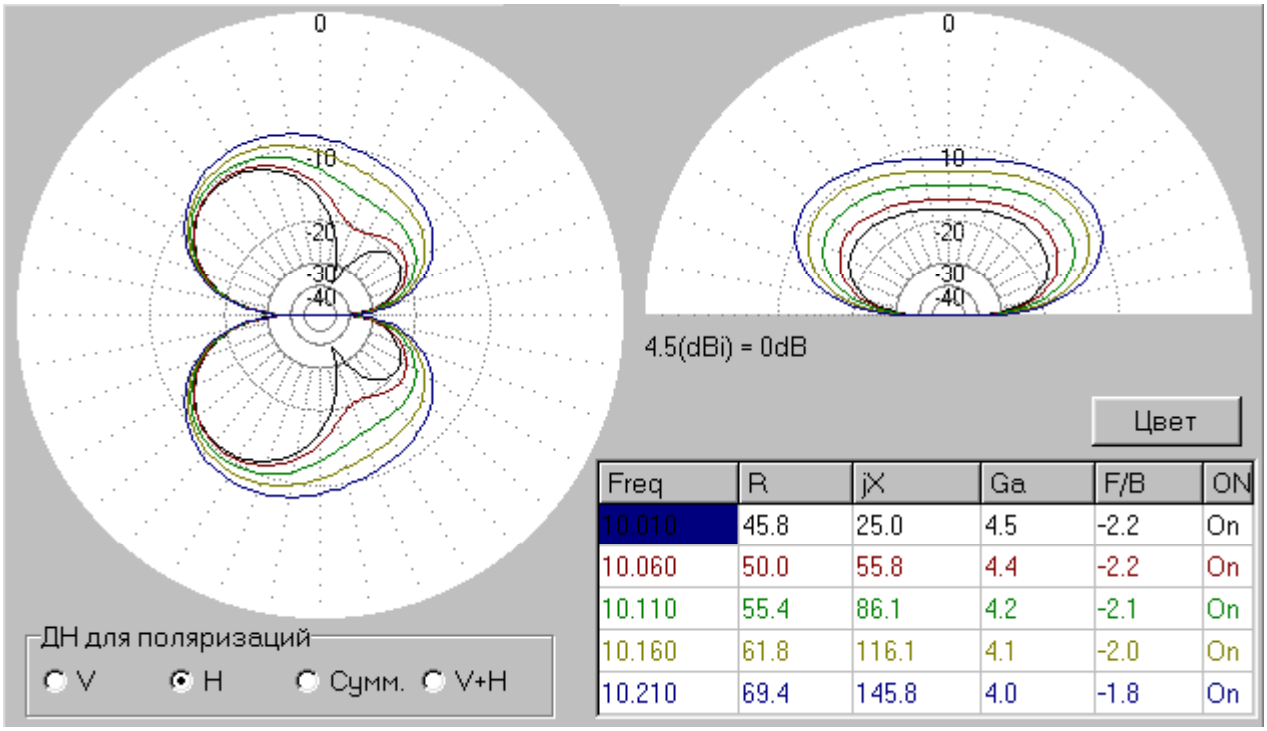
Antenna RV3DA at 30-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
10.010	45.8	25.0	4.5	-2.2	On
10.060	50.0	55.8	4.4	-2.2	On
10.110	55.4	86.1	4.2	-2.1	On
10.160	61.8	116.1	4.1	-2.0	On
10.210	69.4	145.8	4.0	-1.8	On

Horizon Radiation Pattern

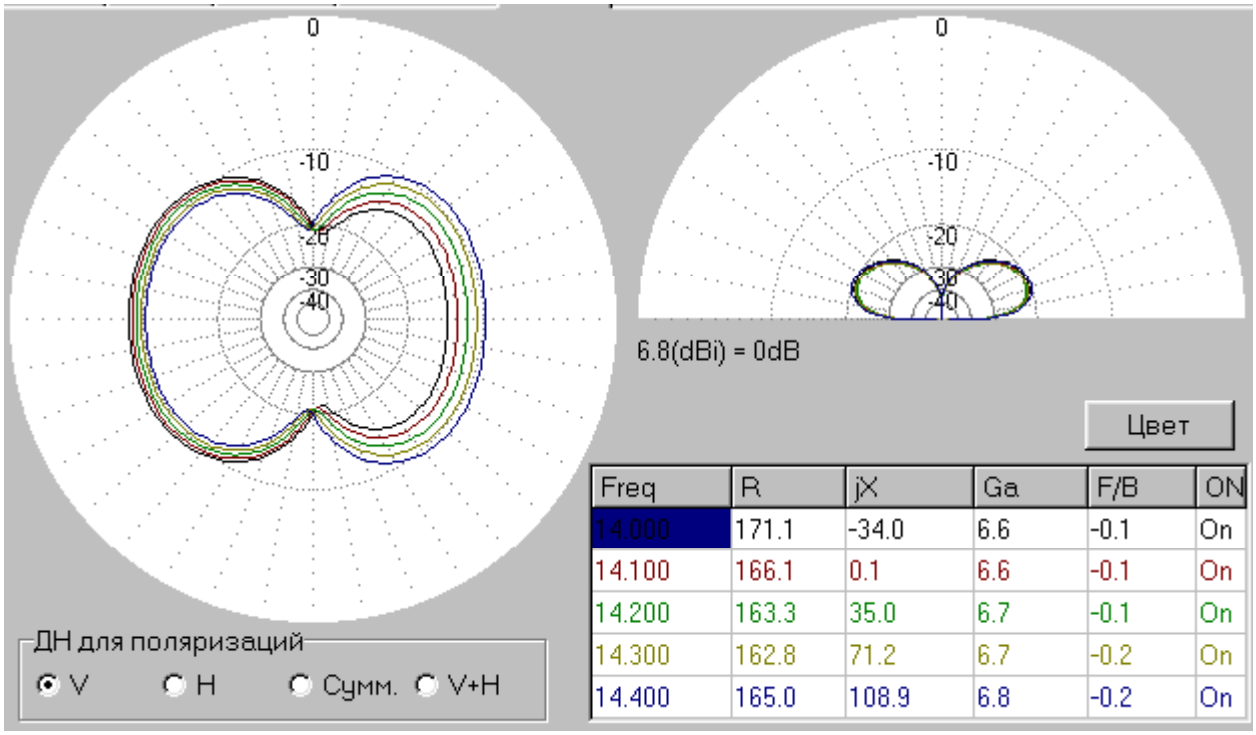


Freq	R	jX	Ga	F/B	ON
10.010	45.8	25.0	4.5	-2.2	On
10.060	50.0	55.8	4.4	-2.2	On
10.110	55.4	86.1	4.2	-2.1	On
10.160	61.8	116.1	4.1	-2.0	On
10.210	69.4	145.8	4.0	-1.8	On

Comments: Antenna has strong vertical radiation. Not bad pattern in the vertical plane. Antenna has strong radiation at high corners that allows to do local QSOs. Antenna has almost circular pattern in horizontal plane. Antenna provides DX and local QSOs. Antenna has Z= 55+ j86-Ohms at 1010-kHz. Any ATU does good matching for such load.

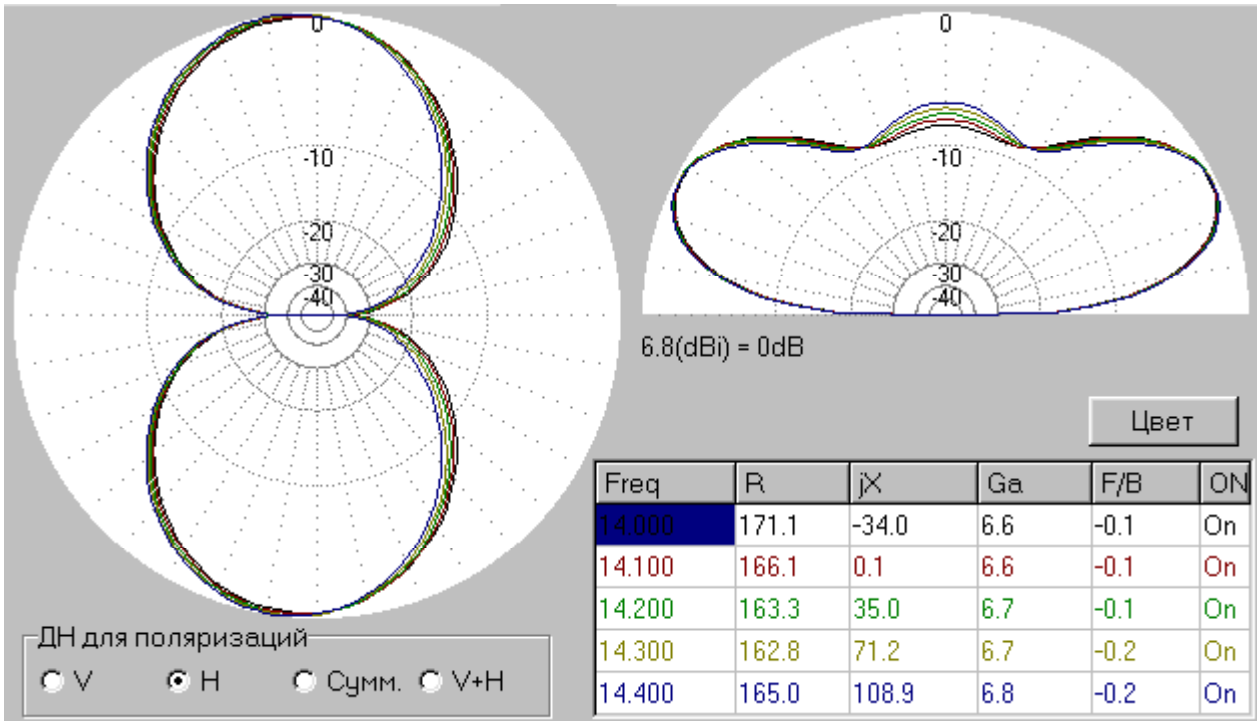
Antenna RV3DA at 20-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
14.000	171.1	-34.0	6.6	-0.1	On
14.100	166.1	0.1	6.6	-0.1	On
14.200	163.3	35.0	6.7	-0.1	On
14.300	162.8	71.2	6.7	-0.2	On
14.400	165.0	108.9	6.8	-0.2	On

Horizon Radiation Pattern

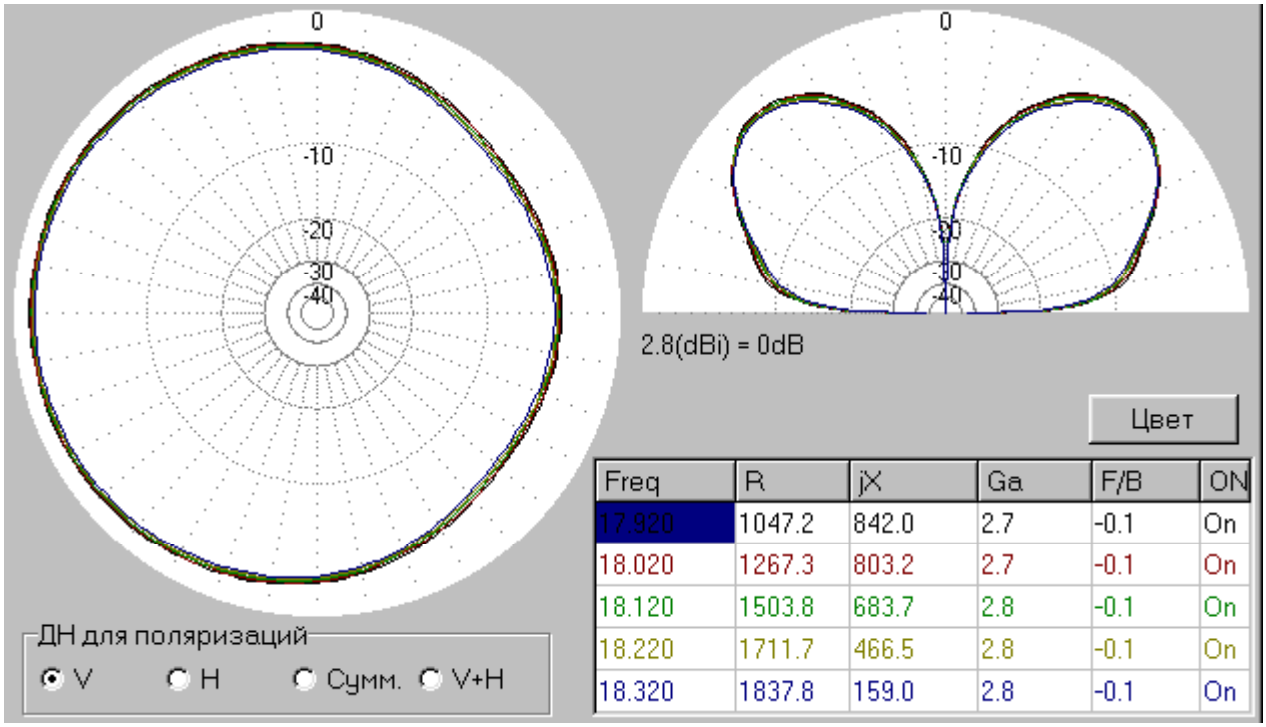


Freq	R	jX	Ga	F/B	ON
14.000	171.1	-34.0	6.6	-0.1	On
14.100	166.1	0.1	6.6	-0.1	On
14.200	163.3	35.0	6.7	-0.1	On
14.300	162.8	71.2	6.7	-0.2	On
14.400	165.0	108.9	6.8	-0.2	On

Comments: Antenna has strong horizon radiation. A good pattern in the vertical plane. Antenna has a strong radiation at low corners that allows to do DX QSOs. Antenna has almost “eight- figure” pattern in horizontal plane, so, it demands to choose a proper direction before an installation of the antenna. Antenna provides DX QSOs. Antenna has the resonance at 14100-kHz at 166 Ohms. Any ATU does good matching for the antenna.

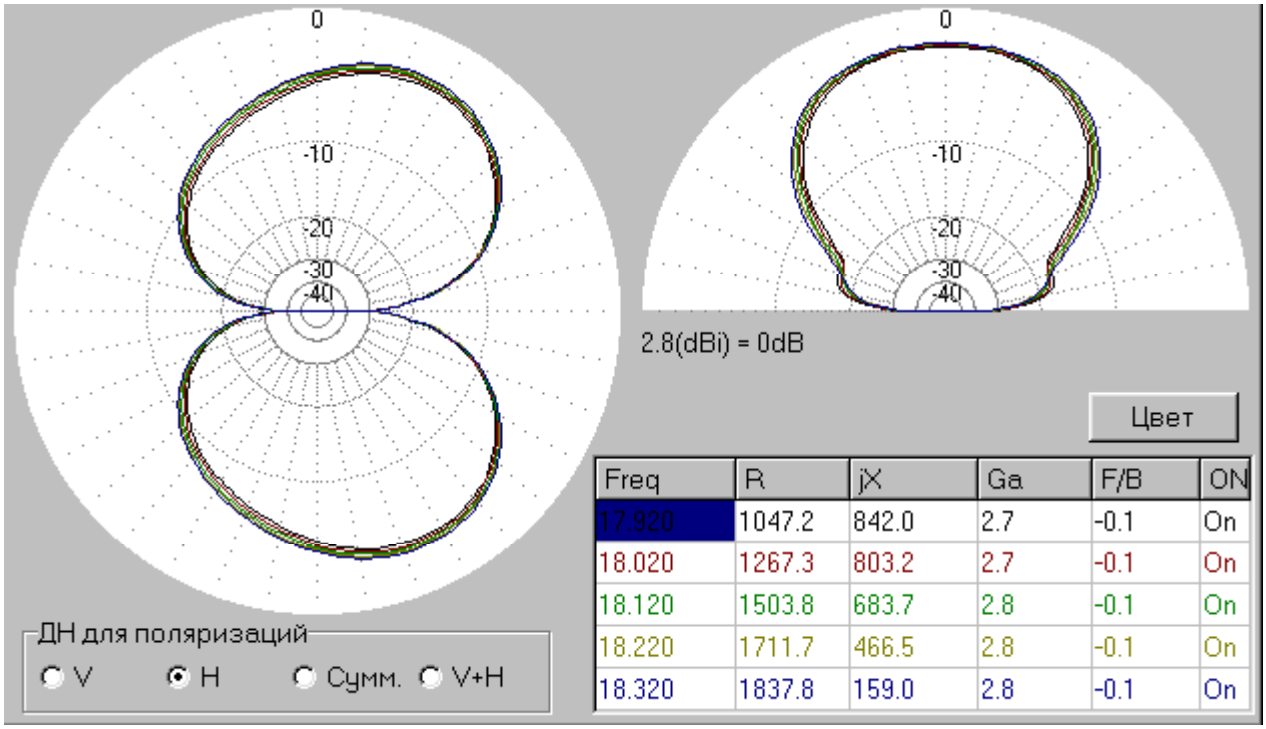
Antenna RV3DA at 17-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
17.920	1047.2	842.0	2.7	-0.1	On
18.020	1267.3	803.2	2.7	-0.1	On
18.120	1503.8	683.7	2.8	-0.1	On
18.220	1711.7	466.5	2.8	-0.1	On
18.320	1837.8	159.0	2.8	-0.1	On

Horizon Radiation Pattern

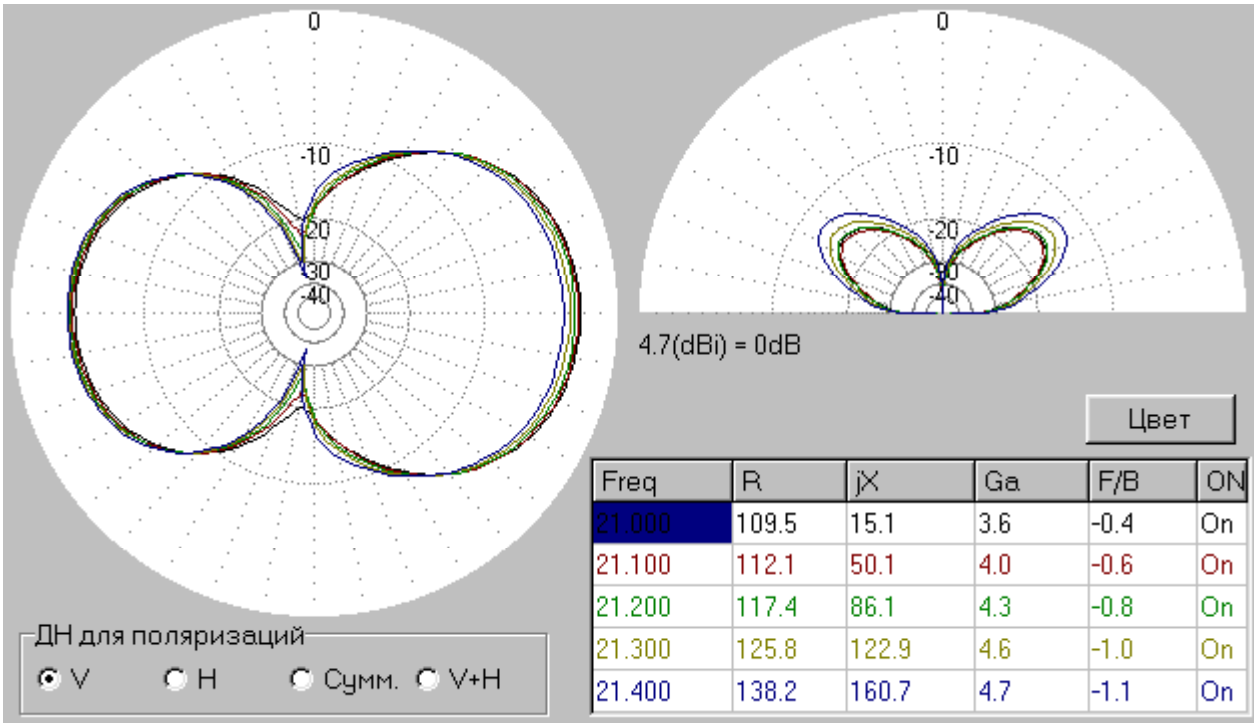


Freq	R	jX	Ga	F/B	ON
17.920	1047.2	842.0	2.7	-0.1	On
18.020	1267.3	803.2	2.7	-0.1	On
18.120	1503.8	683.7	2.8	-0.1	On
18.220	1711.7	466.5	2.8	-0.1	On
18.320	1837.8	159.0	2.8	-0.1	On

Comments: Antenna has both, a strong vertical and strong horizon radiation. For the vertical radiation antenna has a good pattern in the vertical plane, and almost circular pattern in the horizon plane. Pattern for horizon radiation is not so good as to vertical polarization. Antenna can provide DX QSOs at vertical and horizon radiation. Antenna has impedance 1711+j466-Ohms at 18220-kHz. Not all ATUs do good matching for such load.

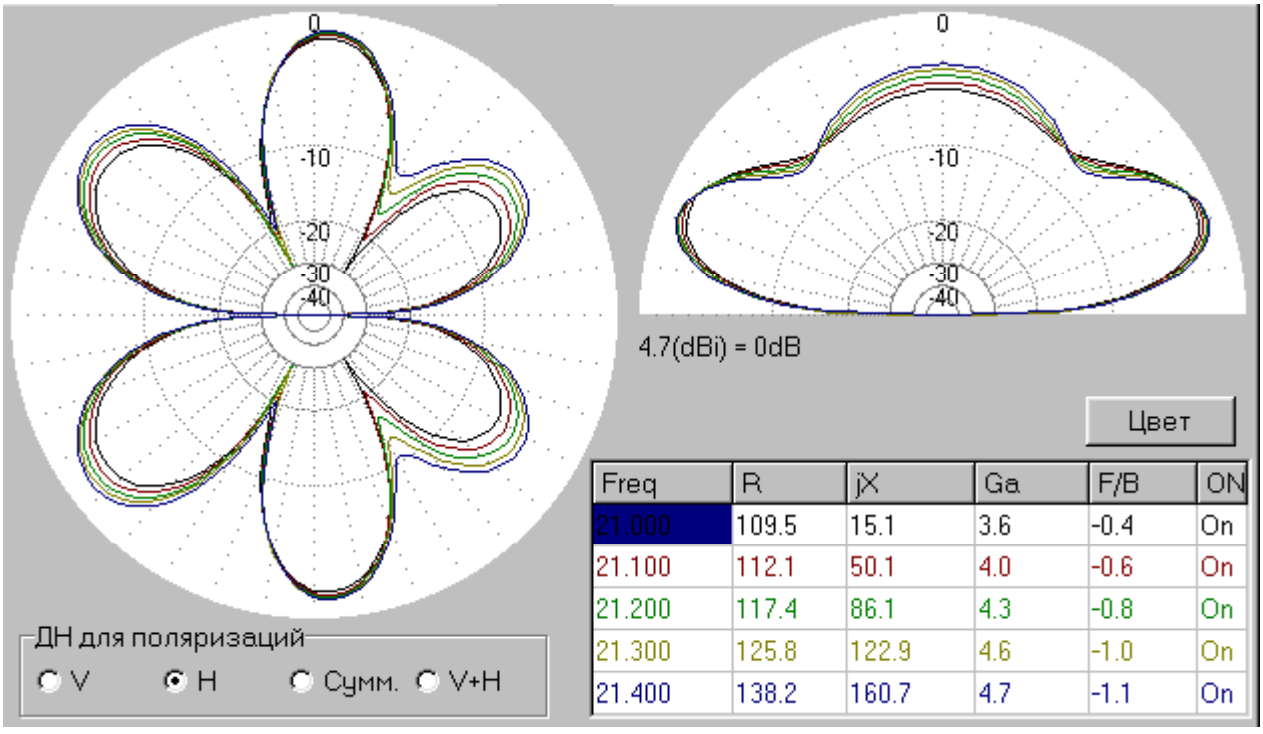
Antenna RV3DA at 15-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
21.000	109.5	15.1	3.6	-0.4	On
21.100	112.1	50.1	4.0	-0.6	On
21.200	117.4	86.1	4.3	-0.8	On
21.300	125.8	122.9	4.6	-1.0	On
21.400	138.2	160.7	4.7	-1.1	On

Horizon Radiation Pattern

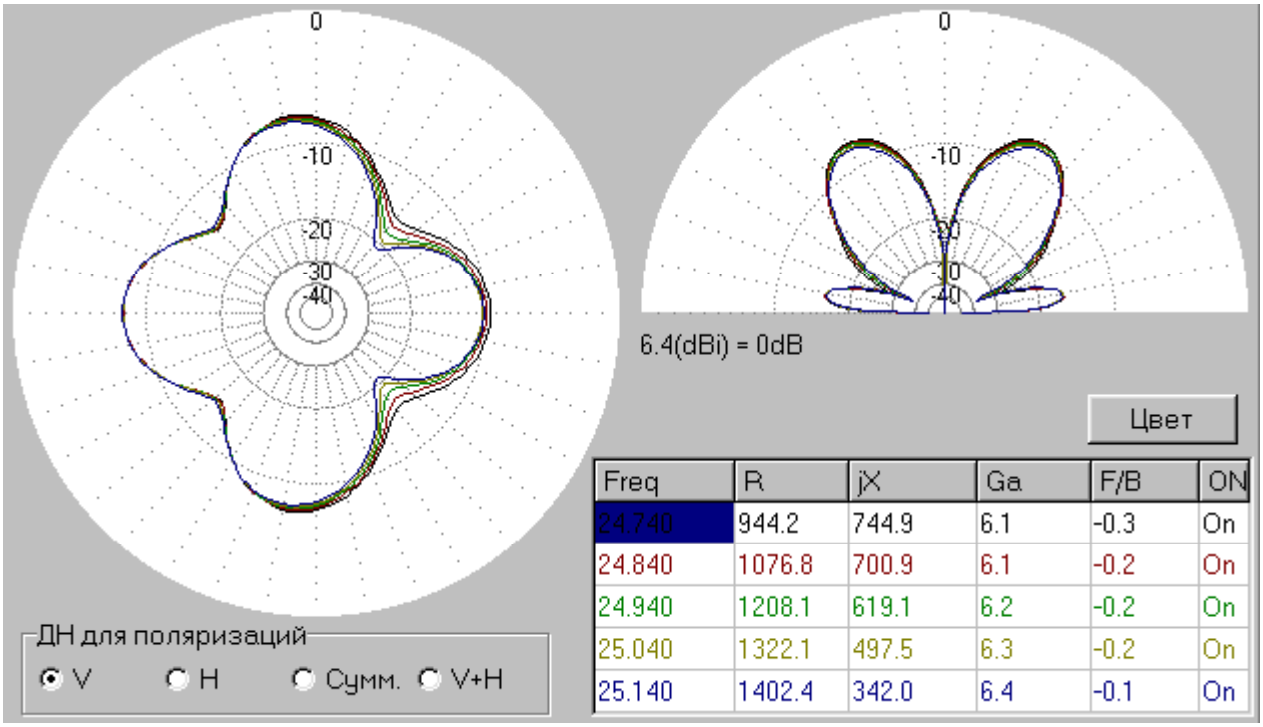


Freq	R	jX	Ga	F/B	ON
21.000	109.5	15.1	3.6	-0.4	On
21.100	112.1	50.1	4.0	-0.6	On
21.200	117.4	86.1	4.3	-0.8	On
21.300	125.8	122.9	4.6	-1.0	On
21.400	138.2	160.7	4.7	-1.1	On

Comments: Antenna has both, a strong vertical and strong horizon radiation. For the horizon radiation antenna has a good pattern in the vertical plane with low lobes, and six-lobes pattern in the horizon plane. Pattern for the horizon radiation is not so good as to vertical polarization. Antenna can provide DX QSOs at vertical radiation. Antenna has impedance 117+j86-Ohms at 21200-kHz. Any ATU makes good matching for such load.

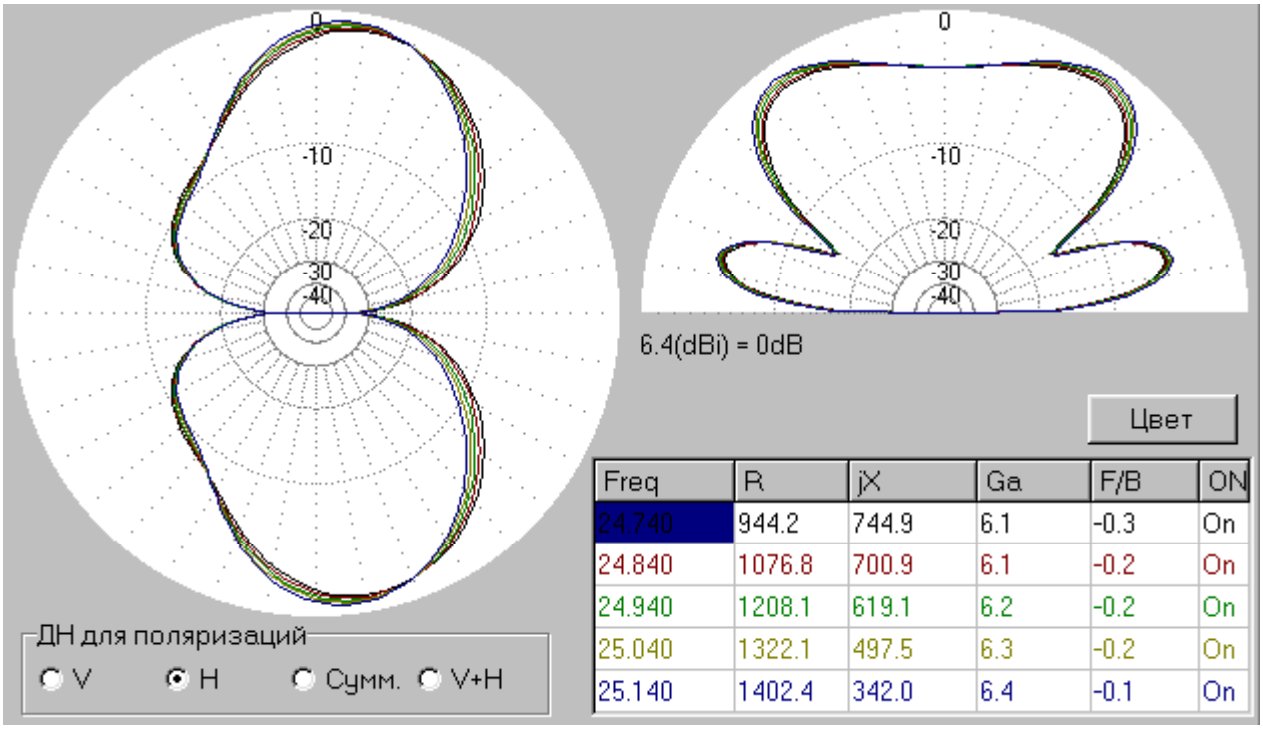
Antenna RV3DA at 12-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
24.740	944.2	744.9	6.1	-0.3	On
24.840	1076.8	700.9	6.1	-0.2	On
24.940	1208.1	619.1	6.2	-0.2	On
25.040	1322.1	497.5	6.3	-0.2	On
25.140	1402.4	342.0	6.4	-0.1	On

Horizon Radiation Pattern

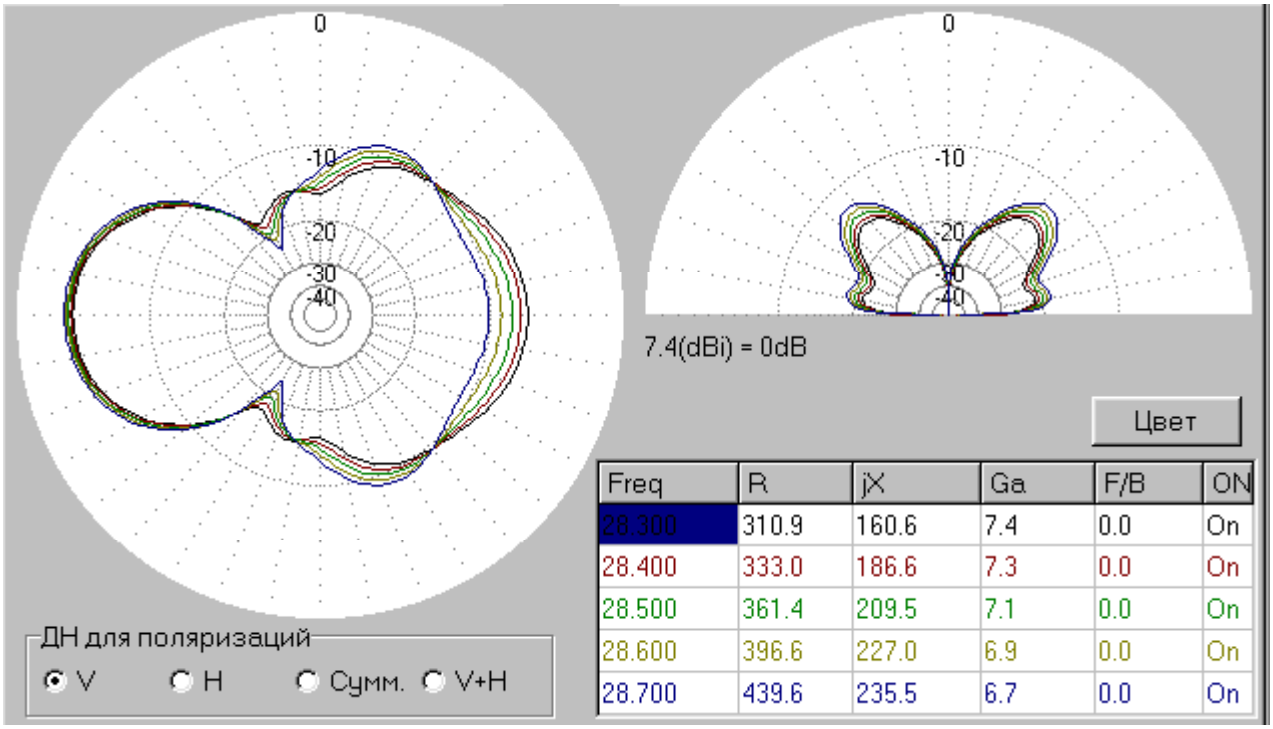


Freq	R	jX	Ga	F/B	ON
24.740	944.2	744.9	6.1	-0.3	On
24.840	1076.8	700.9	6.1	-0.2	On
24.940	1208.1	619.1	6.2	-0.2	On
25.040	1322.1	497.5	6.3	-0.2	On
25.140	1402.4	342.0	6.4	-0.1	On

Comments: Antenna has strong horizon radiation. For the horizon radiation antenna has a not bad pattern in the vertical plane, and “eight- figure” pattern in the horizon plane. Antenna can provide DX QSOs at horizon radiation. Antenna has impedance 1208+j619-Ohms at 21200-kHz. Not all ATUs do good matching for such load.

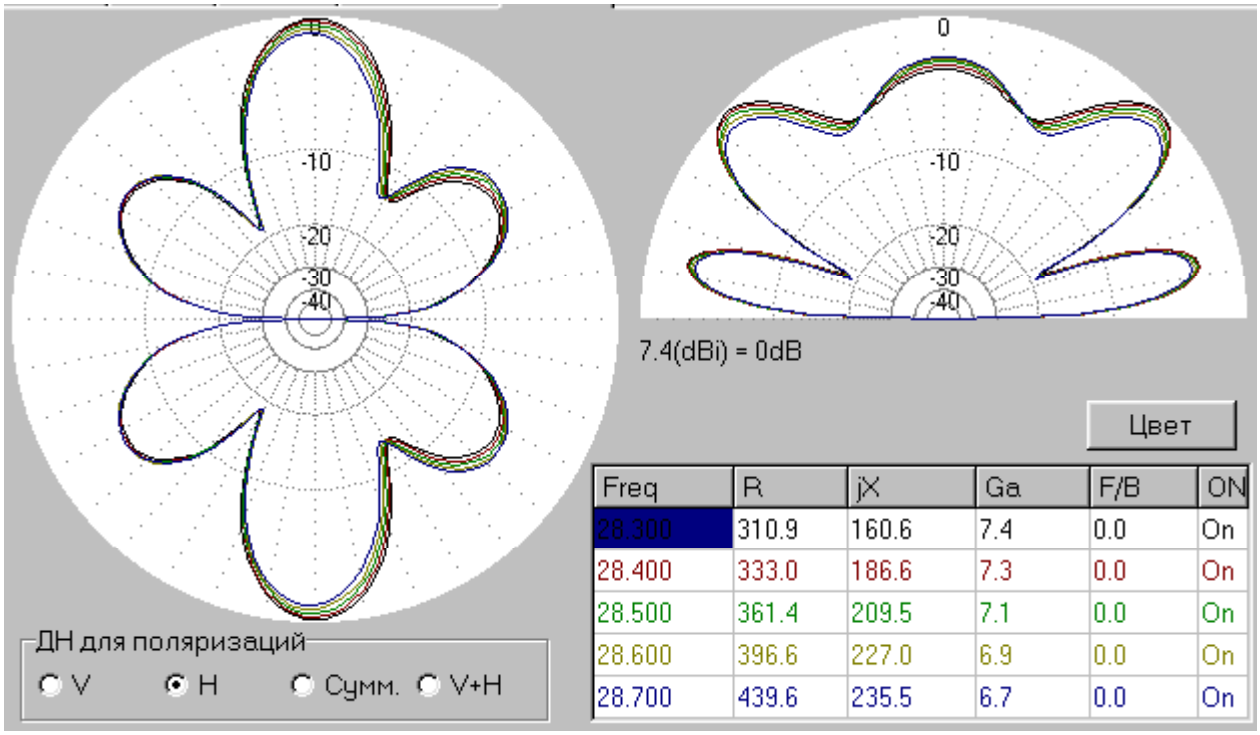
Antenna RV3DA at 10-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
28.300	310.9	160.6	7.4	0.0	On
28.400	333.0	186.6	7.3	0.0	On
28.500	361.4	209.5	7.1	0.0	On
28.600	396.6	227.0	6.9	0.0	On
28.700	439.6	235.5	6.7	0.0	On

Horizon Radiation Pattern

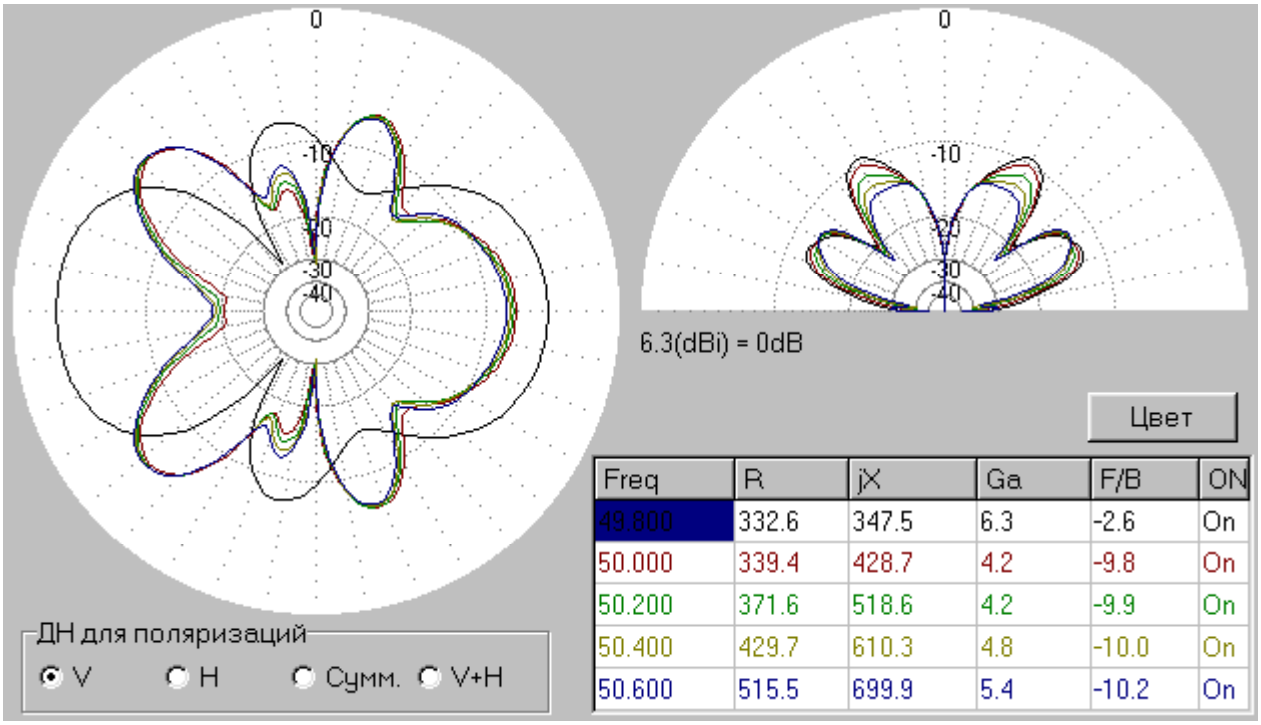


Freq	R	jX	Ga	F/B	ON
28.300	310.9	160.6	7.4	0.0	On
28.400	333.0	186.6	7.3	0.0	On
28.500	361.4	209.5	7.1	0.0	On
28.600	396.6	227.0	6.9	0.0	On
28.700	439.6	235.5	6.7	0.0	On

Comments: Antenna has strong horizon radiation. For the horizon radiation antenna has a not bad pattern in the vertical plane, and “eight- figure” pattern dropped to lobes in the horizon plane. Antenna can provide DX QSOs at the horizon radiation. Antenna has impedance 396+j227-Ohms at 28600-kHz. Not all ATUs do good matching for such load.

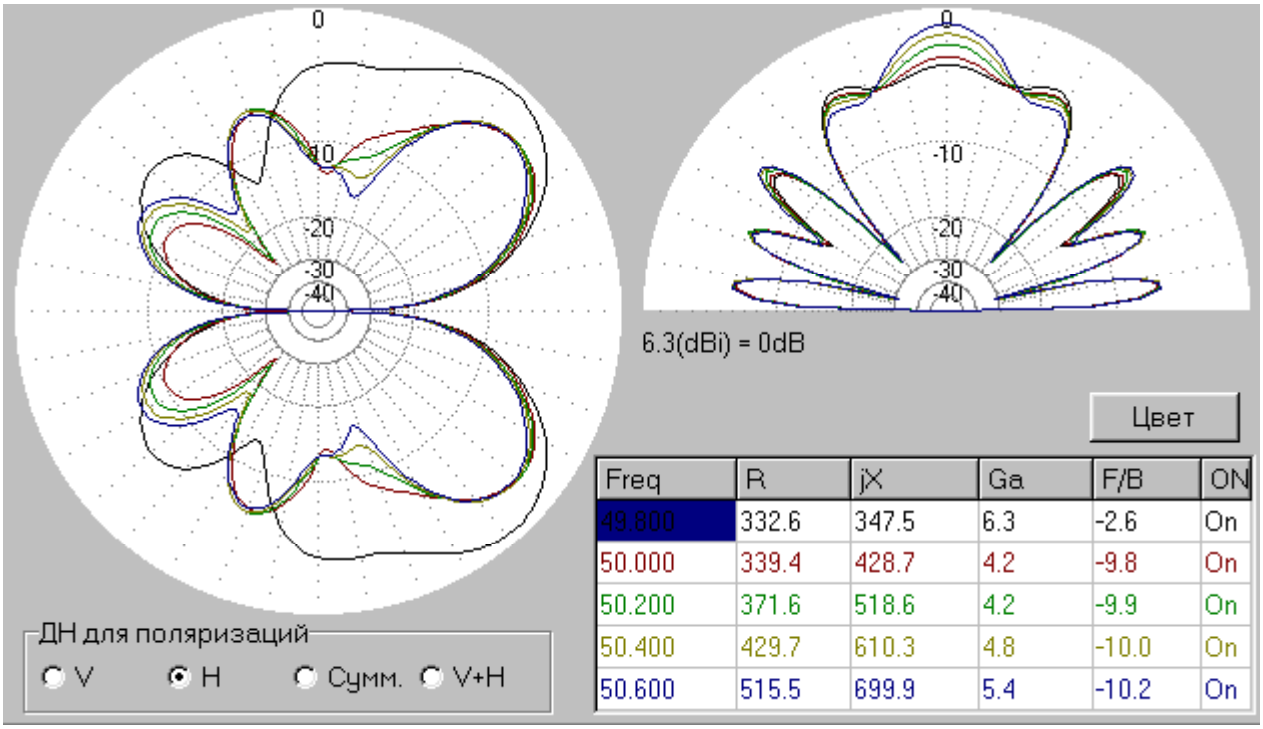
Antenna RV3DA at 6-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
49.800	332.6	347.5	6.3	-2.6	On
50.000	339.4	428.7	4.2	-9.8	On
50.200	371.6	518.6	4.2	-9.9	On
50.400	429.7	610.3	4.8	-10.0	On
50.600	515.5	699.9	5.4	-10.2	On

Horizon Radiation Pattern



Freq	R	jX	Ga	F/B	ON
49.800	332.6	347.5	6.3	-2.6	On
50.000	339.4	428.7	4.2	-9.8	On
50.200	371.6	518.6	4.2	-9.9	On
50.400	429.7	610.3	4.8	-10.0	On
50.600	515.5	699.9	5.4	-10.2	On

Comments: Antenna has horizon and vertical radiation. For the horizon radiation antenna has a not bad pattern in the vertical plane, and “eight- figure” pattern dropped to lobes in the horizon plane. Antenna can provide DX QSOs at horizon radiation. Antenna has impedance 429+j599-Ohms at 50600-kHz. Not all ATUs do good matching for such load.

SHUNT VERTICAL UNIVERSAL HF ANTENNA

Igor Grigorov, Rk3ZK

antentop@mail.ru

*Field universal antenna RV3DA (**see pp.:24- 35 of ANTENTOP- 02- 2004**) works well even at a bad grounding. To hammer into the ground a metal rod in 1 meter length is enough for the grounding. Installation of the antenna takes a little time, it is another its advantage.However, if there is an opportunity to provide a good ground, and there is some free time to spend of for installation of an antenna, it is possible to use a **Shunt Vertical Universal HF Antenna** .*

Figure 1 shows the schematic of the Shunt Vertical Universal HF Antenna. A detailed description of the theory of a Shunt Vertical Universal is given at reference **[1]**.

Apparently, the circuit of the antenna only a bit differs from field universal antenna RV3DA. The differences are: the loop is isolated from the ground, its terminals are shortened, shunts go down from two tops of the triangle loop to the ground. To ground shunts is possible as to universal antenna RV3DA it is done, i.e., a metal rod in 1 meter length is enough for the grounding. Of course, several counterpoises (three and more) in length 0f 5 meters (and more) help to improve the antenna operation. Counterpoises can lay on a surface of the ground.

Shunt Vertical Universal HF Antenna radiates mainly vertical radiation. It is required to use the antenna at woodless surrounding or big losses of high-frequency energy will be. Please, take attention Shunt Vertical Universal HF Antenna has a gain less then universal antenna RV3DA. It is possible to do a design of the Shunt Vertical Universal HF Antenna so, that this one can be easy turned to the field universal antenna RV3DA.

Below given diagram directivity for the antenna obtained with help of free antenna program MMANA (MININEC based). Left diagram is a section of the volumetric diagram directivity of plane X-Y at a zenith corner of the maximum radiation. The right diagram is section of the volumetric diagram directivity of plane X-Z. Also at the right down corner of the pictures is a table with antenna impedance. Please, take attention to the data, you can do decision how you ATU does match of the

Reference:

1. Aizenberg G. Z. Antennas of Short Waves.: Moscow, '
Svyaz", 1985.

73!
Igor Grigorov, RK3ZK

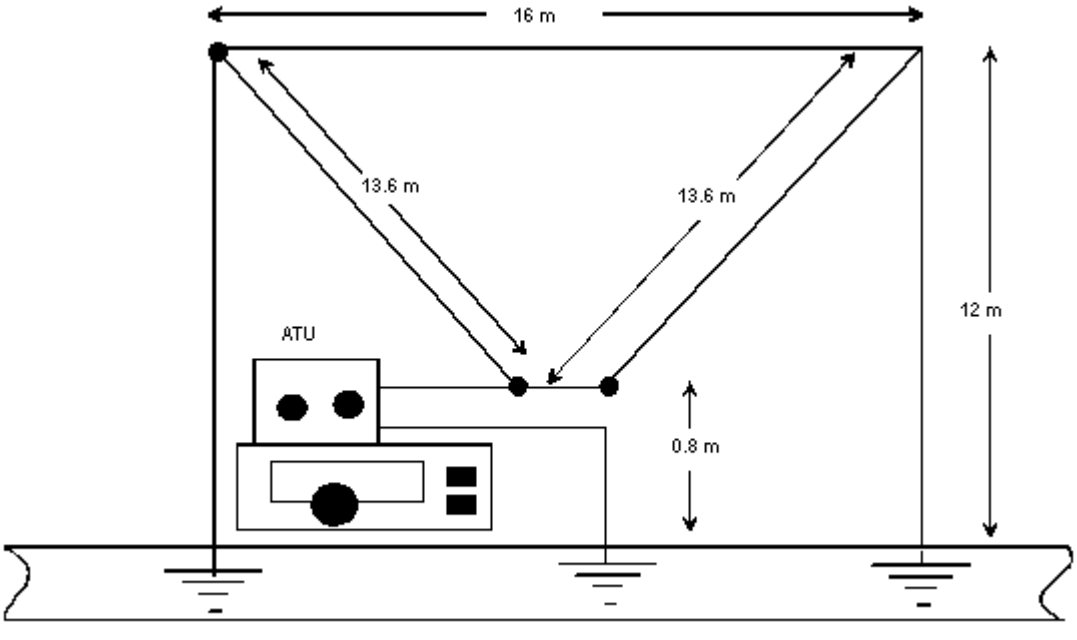
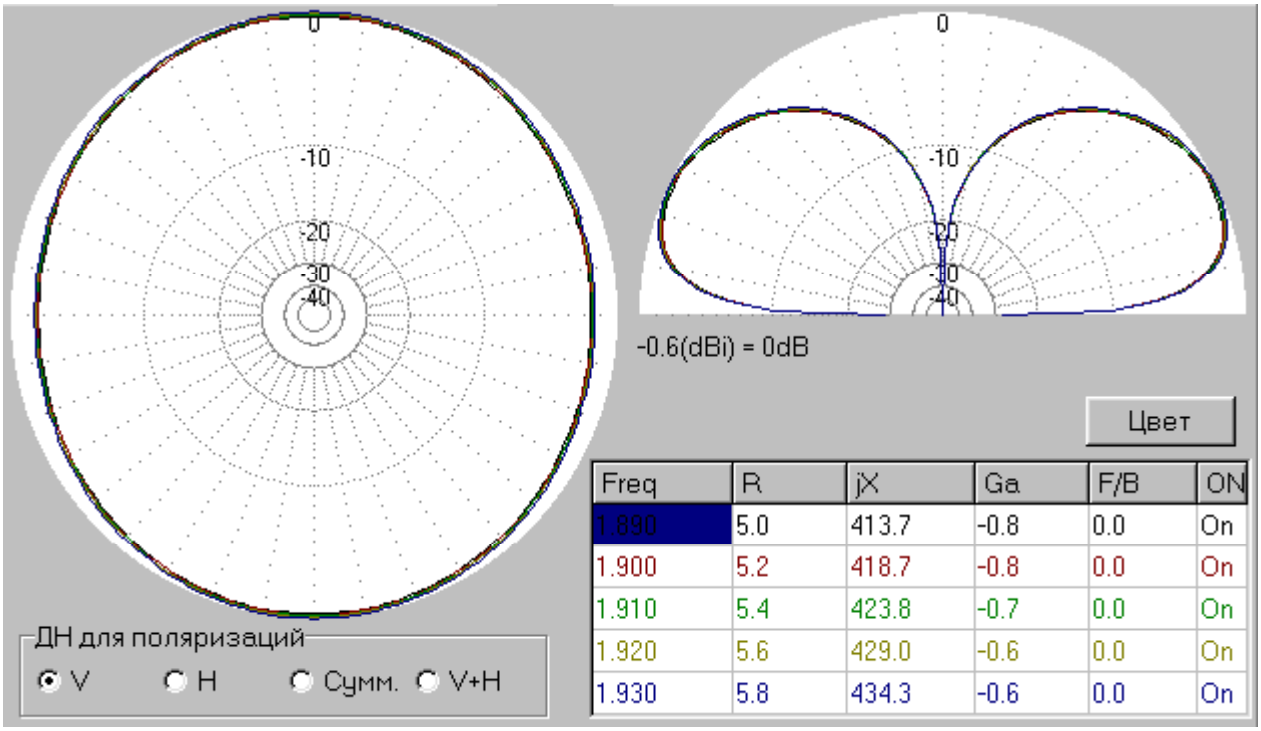


Figure 1

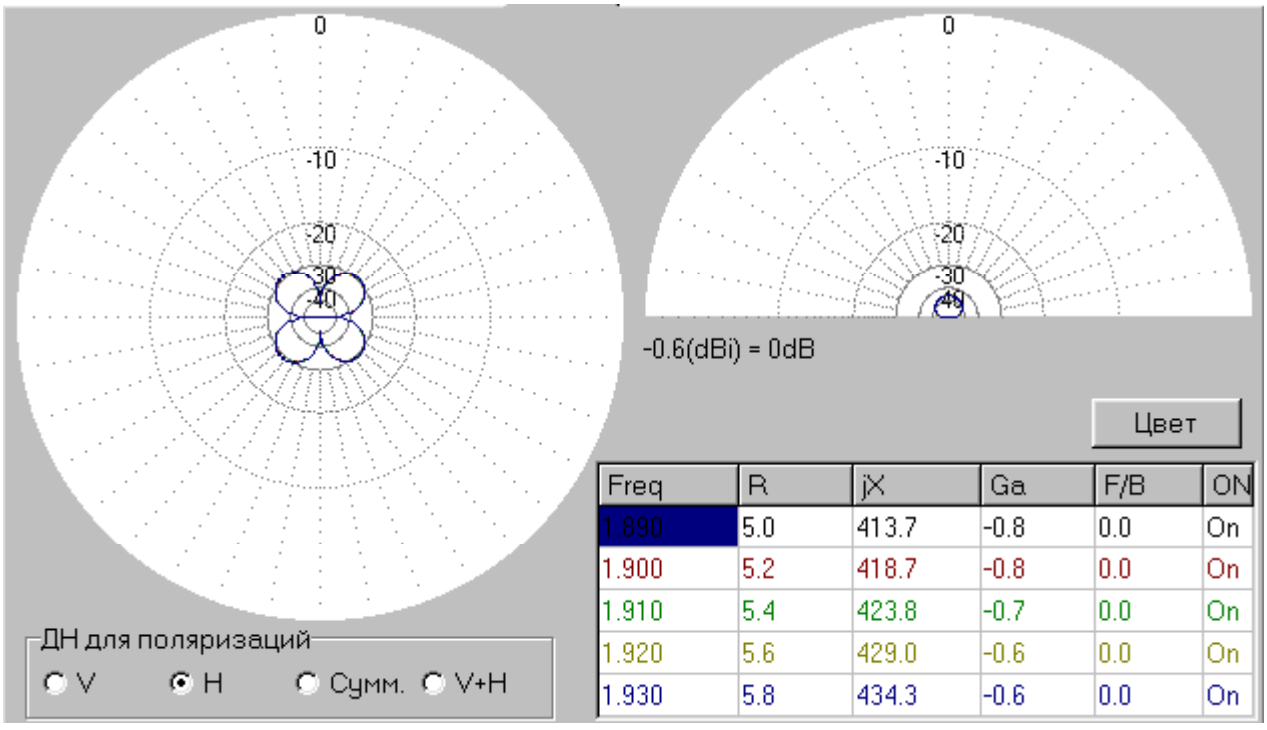
Shunt Vertical Universal HF Antenna at 160-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
1.890	5.0	413.7	-0.8	0.0	On
1.900	5.2	418.7	-0.8	0.0	On
1.910	5.4	423.8	-0.7	0.0	On
1.920	5.6	429.0	-0.6	0.0	On
1.930	5.8	434.3	-0.6	0.0	On

Horizon Radiation Pattern

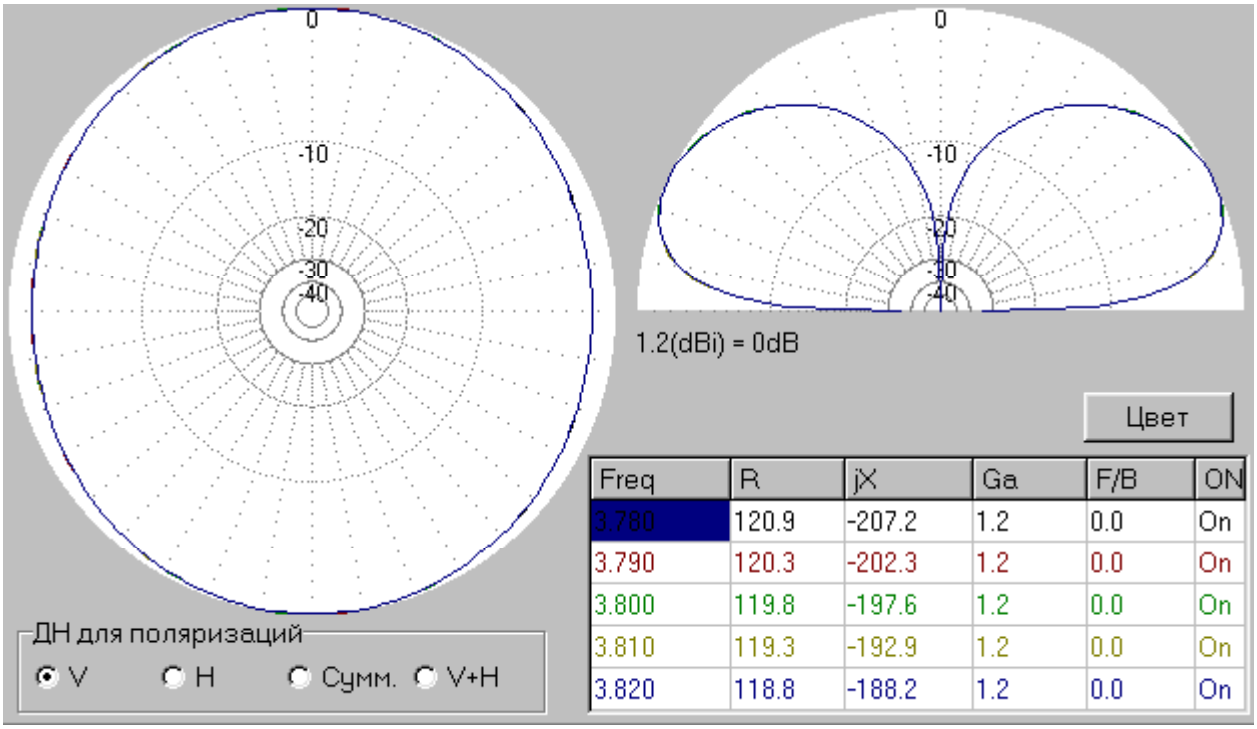


Freq	R	jX	Ga	F/B	ON
1.890	5.0	413.7	-0.8	0.0	On
1.900	5.2	418.7	-0.8	0.0	On
1.910	5.4	423.8	-0.7	0.0	On
1.920	5.6	429.0	-0.6	0.0	On
1.930	5.8	434.3	-0.6	0.0	On

Comments: Antenna radiates radio waves with vertical polarization. A very good pattern with low lobes in the vertical plane. Circular pattern in horizontal plane. It is fine for DX- QSO. But antenna has Z= 5.4+ j423-Ohms at 1910-kHz. Not all ATUs do good matching for such load.

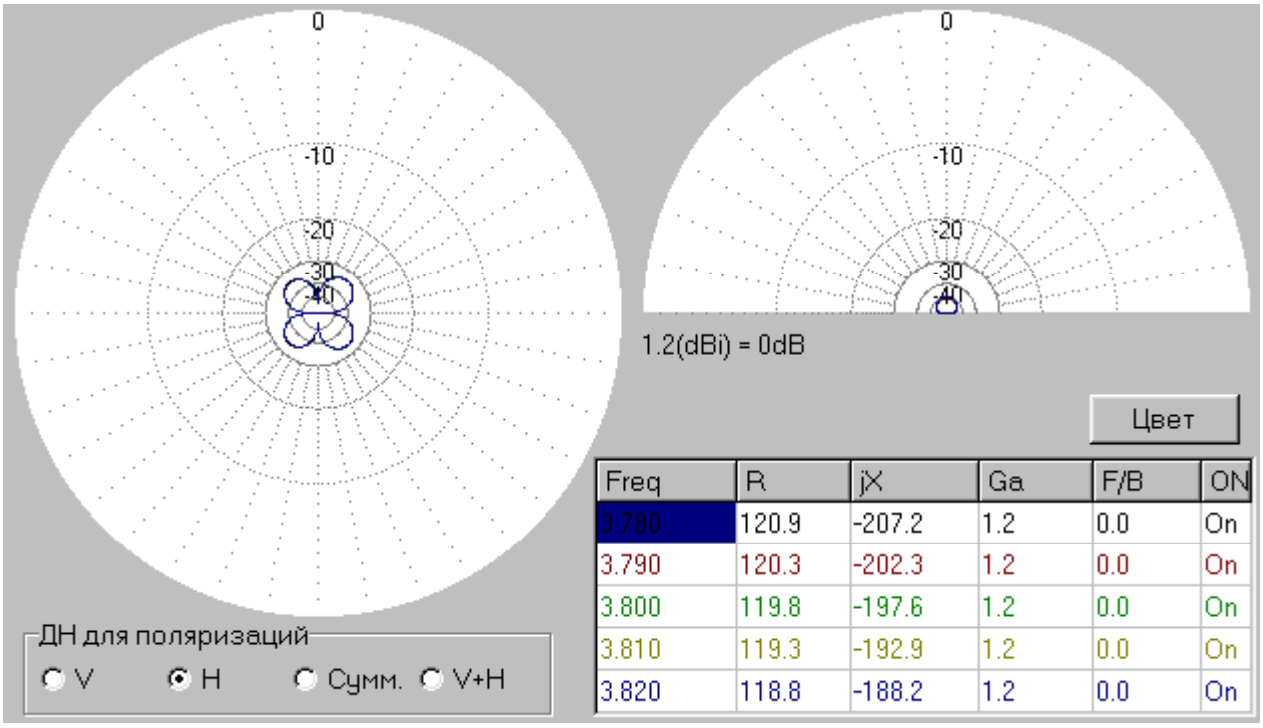
Shunt Vertical Universal HF Antenna at 80-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
3.780	120.9	-207.2	1.2	0.0	On
3.790	120.3	-202.3	1.2	0.0	On
3.800	119.8	-197.6	1.2	0.0	On
3.810	119.3	-192.9	1.2	0.0	On
3.820	118.8	-188.2	1.2	0.0	On

Horizon Radiation Pattern

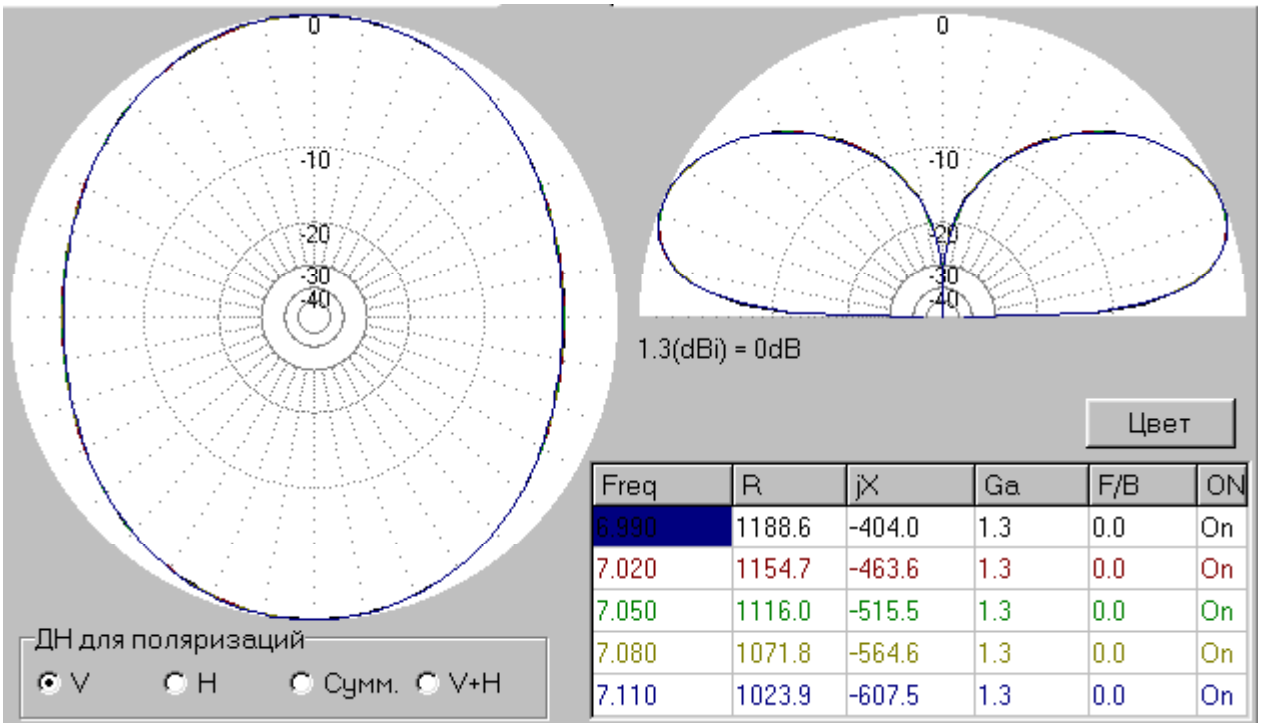


Freq	R	jX	Ga	F/B	ON
3.780	120.9	-207.2	1.2	0.0	On
3.790	120.3	-202.3	1.2	0.0	On
3.800	119.8	-197.6	1.2	0.0	On
3.810	119.3	-192.9	1.2	0.0	On
3.820	118.8	-188.2	1.2	0.0	On

Comments: Antenna radiates radio waves with vertical polarization. A very good pattern with low lobes in the vertical plane. Circular pattern in horizontal plane. It is fine for DX- QSO. Antenna has Z= 119-j197-Ohms at 3800-kHz. Almost any ATU does good matching for such load.

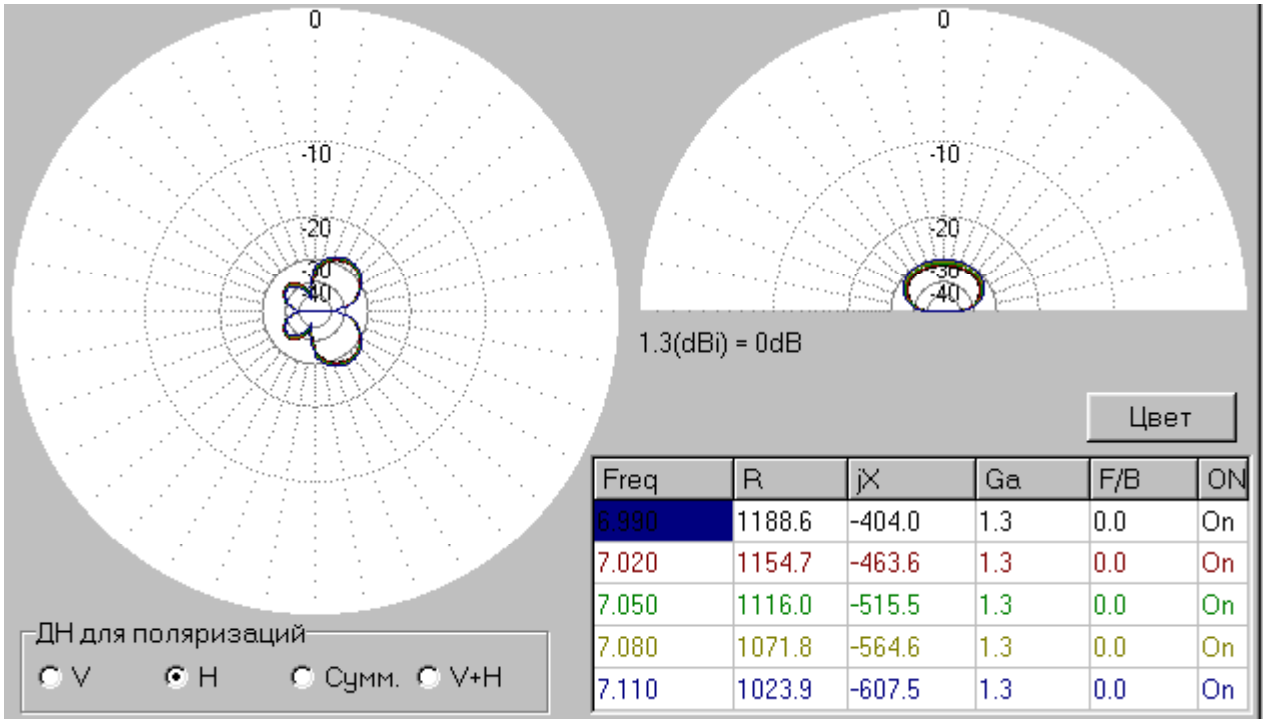
Shunt Vertical Universal HF Antenna at 40-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
6.990	1188.6	-404.0	1.3	0.0	On
7.020	1154.7	-463.6	1.3	0.0	On
7.050	1116.0	-515.5	1.3	0.0	On
7.080	1071.8	-564.6	1.3	0.0	On
7.110	1023.9	-607.5	1.3	0.0	On

Horizon Radiation Pattern

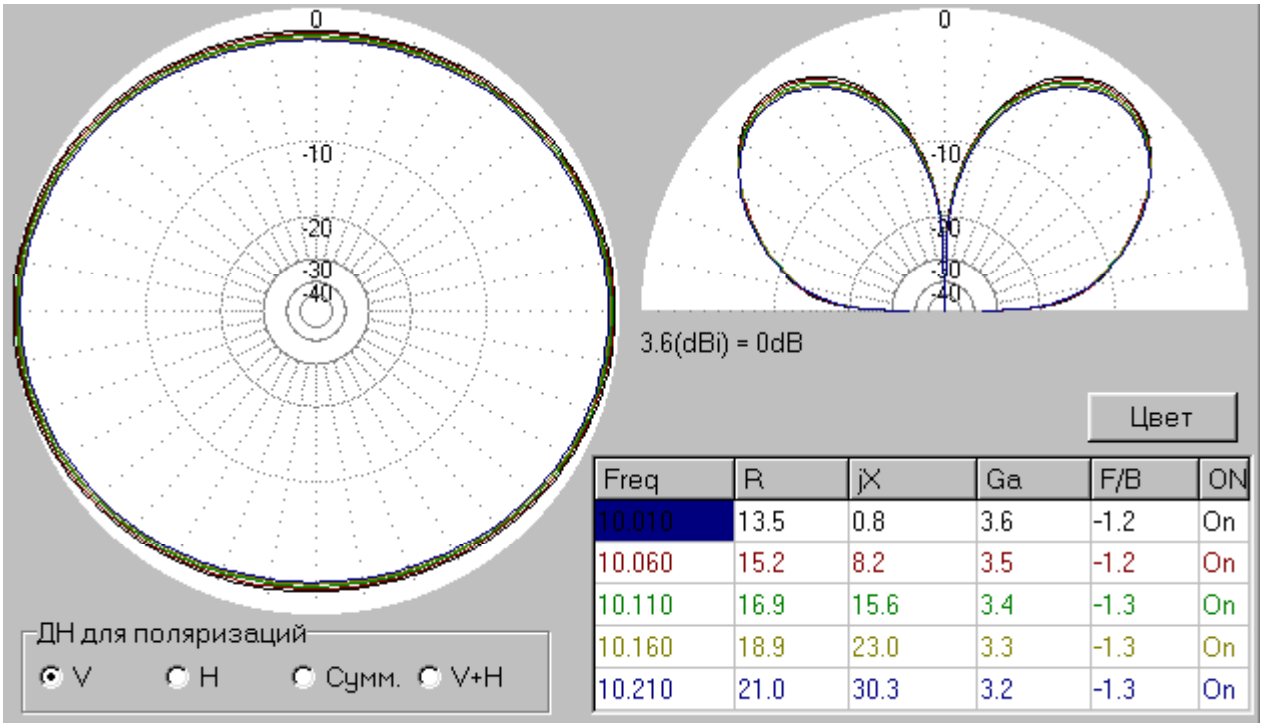


Freq	R	jX	Ga	F/B	ON
6.990	1188.6	-404.0	1.3	0.0	On
7.020	1154.7	-463.6	1.3	0.0	On
7.050	1116.0	-515.5	1.3	0.0	On
7.080	1071.8	-564.6	1.3	0.0	On
7.110	1023.9	-607.5	1.3	0.0	On

Comments: Antenna radiates radio waves with vertical polarization. A very good pattern with low lobes in the vertical plane. Almost a circular pattern in horizontal plane. It is fine for DX- QSO. Antenna has Z=1116- j515-Ohms at 7050-kHz. Not all ATUs do good matching for such load.

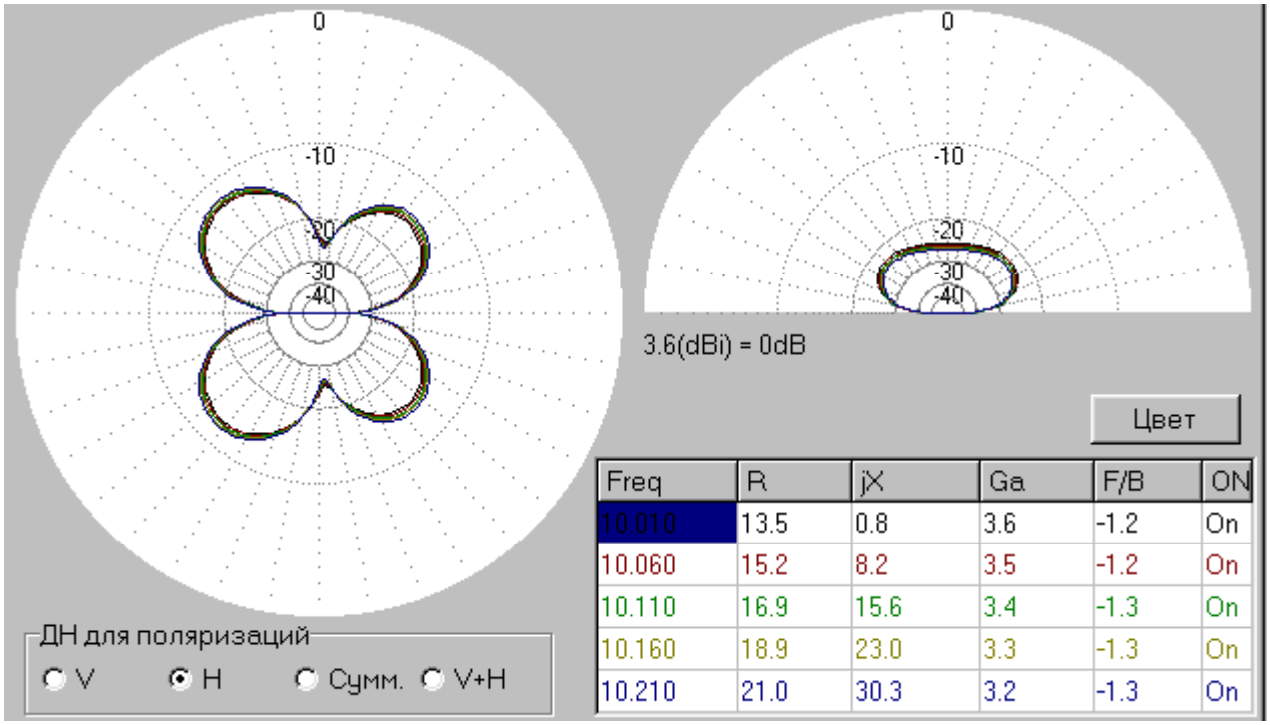
Shunt Vertical Universal HF Antenna at 30-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
10.010	13.5	0.8	3.6	-1.2	On
10.060	15.2	8.2	3.5	-1.2	On
10.110	16.9	15.6	3.4	-1.3	On
10.160	18.9	23.0	3.3	-1.3	On
10.210	21.0	30.3	3.2	-1.3	On

Horizon Radiation Pattern

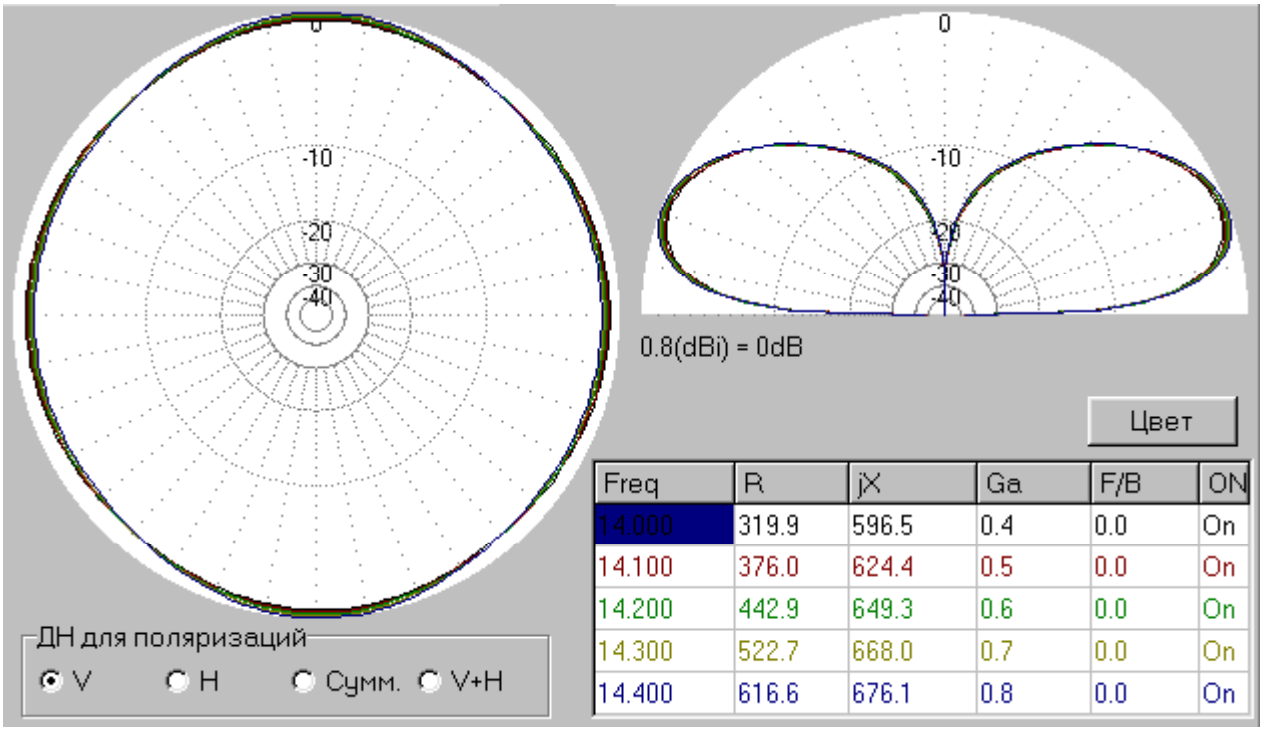


Freq	R	jX	Ga	F/B	ON
10.010	13.5	0.8	3.6	-1.2	On
10.060	15.2	8.2	3.5	-1.2	On
10.110	16.9	15.6	3.4	-1.3	On
10.160	18.9	23.0	3.3	-1.3	On
10.210	21.0	30.3	3.2	-1.3	On

Comments: Antenna has strong vertical radiation. A good pattern in the vertical plane. Antenna has strong radiation at high corners that allows to do local QSOs. Antenna has almost circular pattern in horizontal plane. Antenna provides DX and local QSOs. Antenna has Z= 16.9+ j15.6-Ohms at 10110-kHz. Any ATU does good matching for such load.

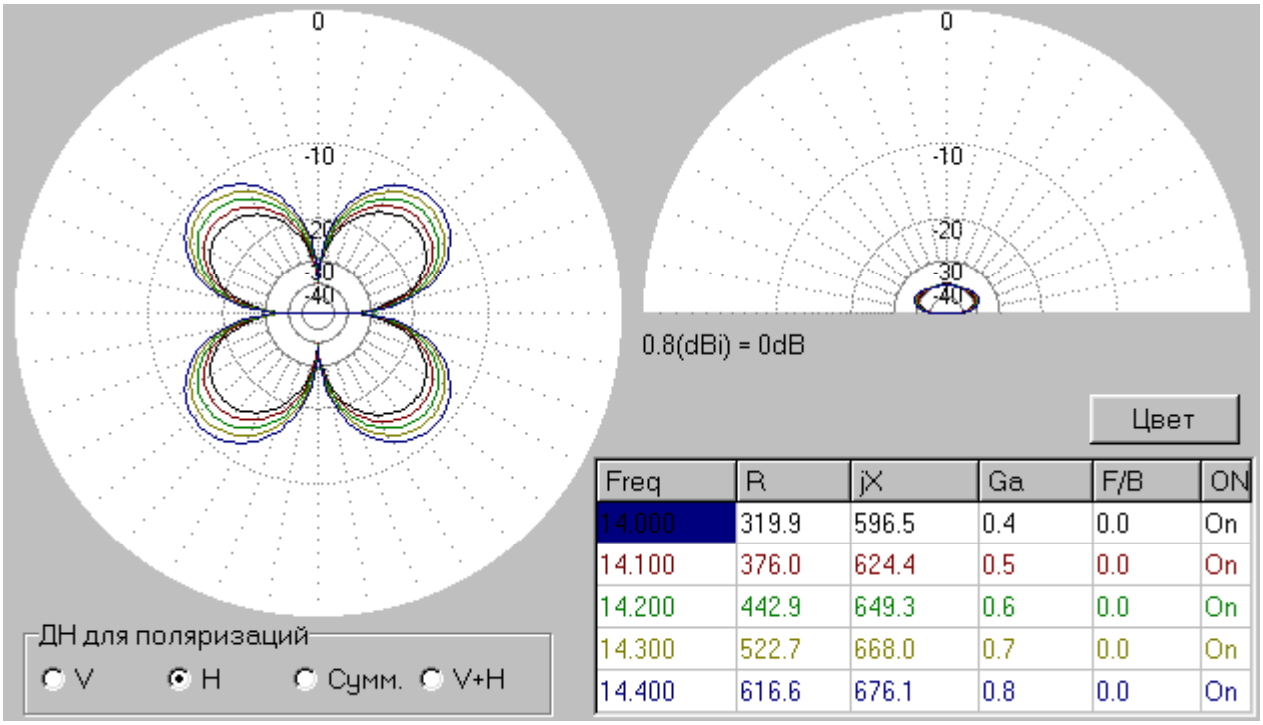
Shunt Vertical Universal HF Antenna at 20-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
14.000	319.9	596.5	0.4	0.0	On
14.100	376.0	624.4	0.5	0.0	On
14.200	442.9	649.3	0.6	0.0	On
14.300	522.7	668.0	0.7	0.0	On
14.400	616.6	676.1	0.8	0.0	On

Horizon Radiation Pattern

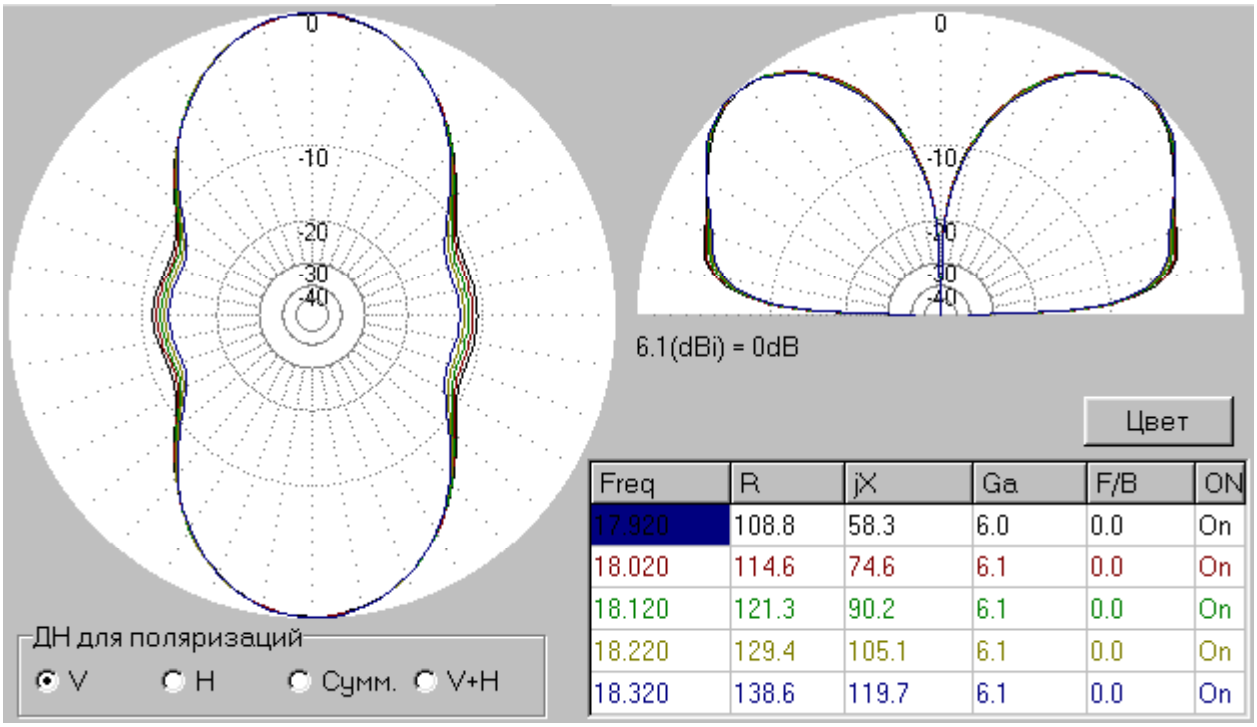


Freq	R	jX	Ga	F/B	ON
14.000	319.9	596.5	0.4	0.0	On
14.100	376.0	624.4	0.5	0.0	On
14.200	442.9	649.3	0.6	0.0	On
14.300	522.7	668.0	0.7	0.0	On
14.400	616.6	676.1	0.8	0.0	On

Comments: Antenna has strong vertical radiation. A very good pattern in the vertical plane. Antenna has almost circular pattern in horizontal plane. Antenna provides DX QSOs. Antenna has Z= 442+ j649-Ohms at 14200-kHz. Not all ATUs do good matching for such load.

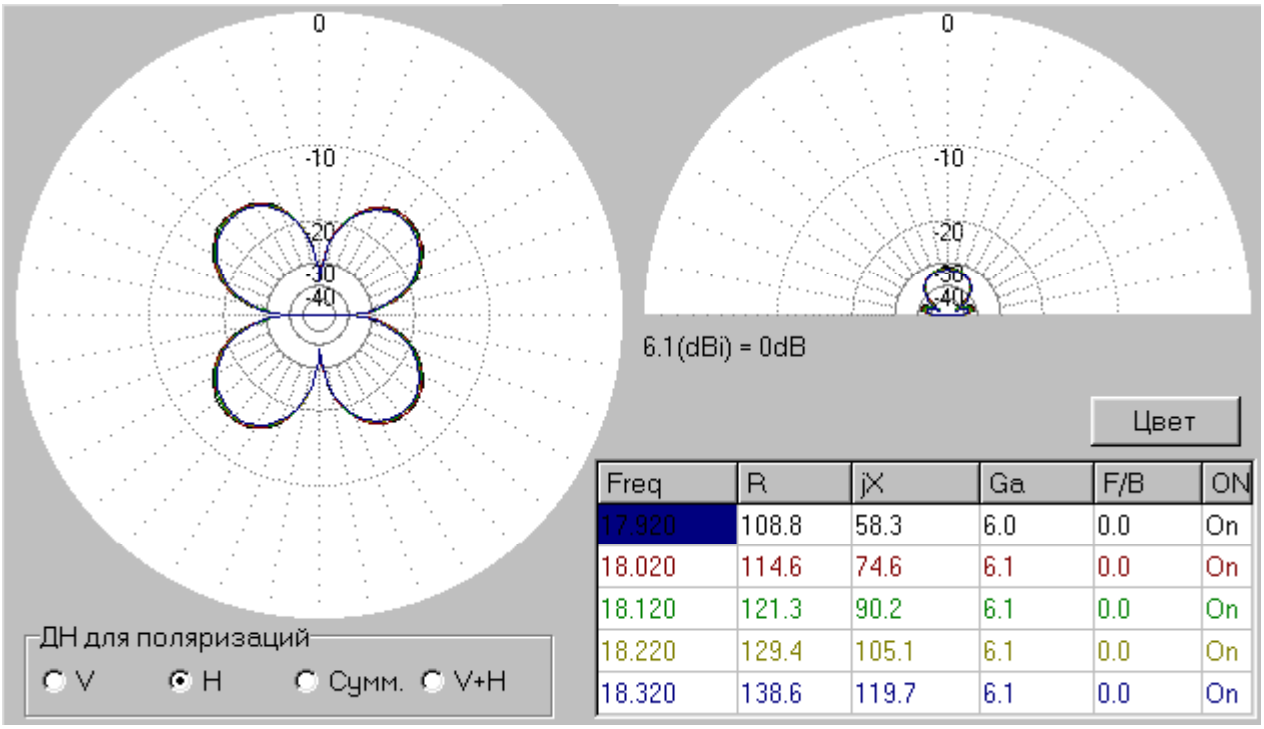
Shunt Vertical Universal HF Antenna at 17-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
17.920	108.8	58.3	6.0	0.0	On
18.020	114.6	74.6	6.1	0.0	On
18.120	121.3	90.2	6.1	0.0	On
18.220	129.4	105.1	6.1	0.0	On
18.320	138.6	119.7	6.1	0.0	On

Horizon Radiation Pattern

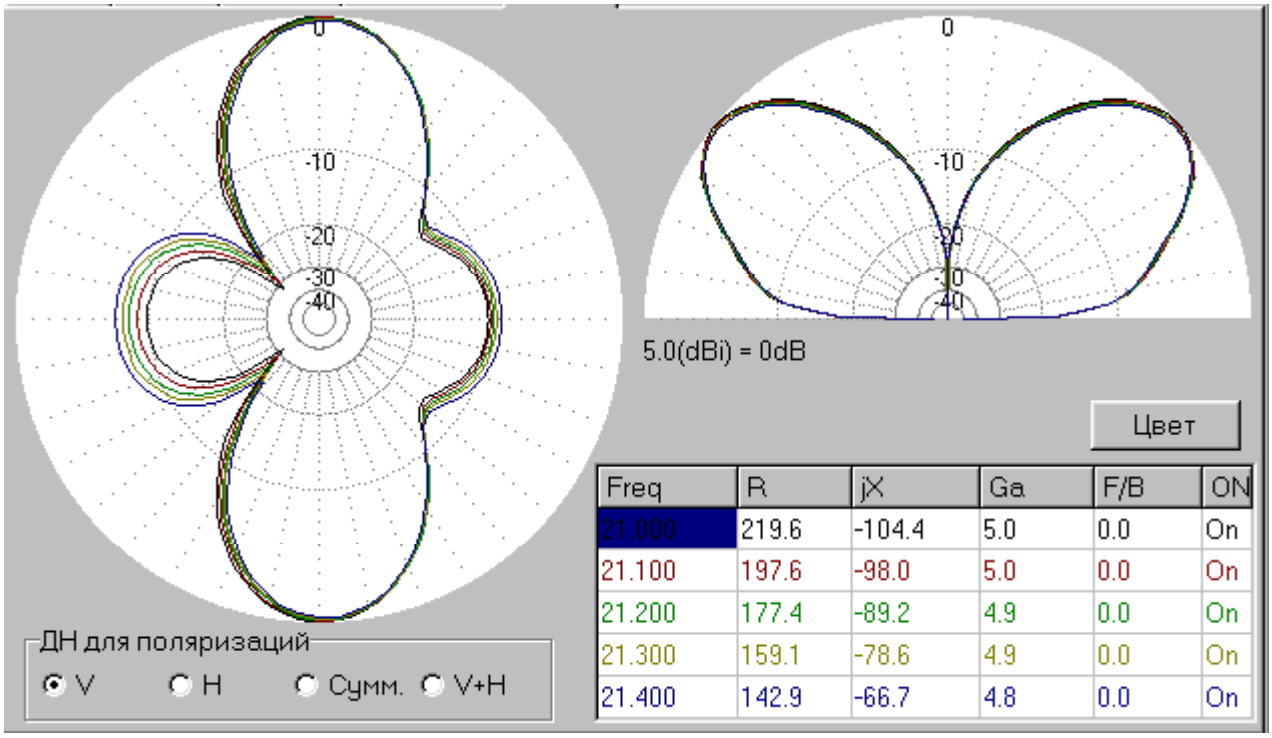


Freq	R	jX	Ga	F/B	ON
17.920	108.8	58.3	6.0	0.0	On
18.020	114.6	74.6	6.1	0.0	On
18.120	121.3	90.2	6.1	0.0	On
18.220	129.4	105.1	6.1	0.0	On
18.320	138.6	119.7	6.1	0.0	On

Comments: Antenna has strong vertical radiation. A good pattern in the vertical plane. Antenna has egg shape pattern in horizontal plane, so, it demands to choose a proper direction before an installation of the antenna. Antenna provides DX QSOs. Antenna has Z= 121+ j90-Ohms at 18120-kHz. Any ATU does good matching for such load.

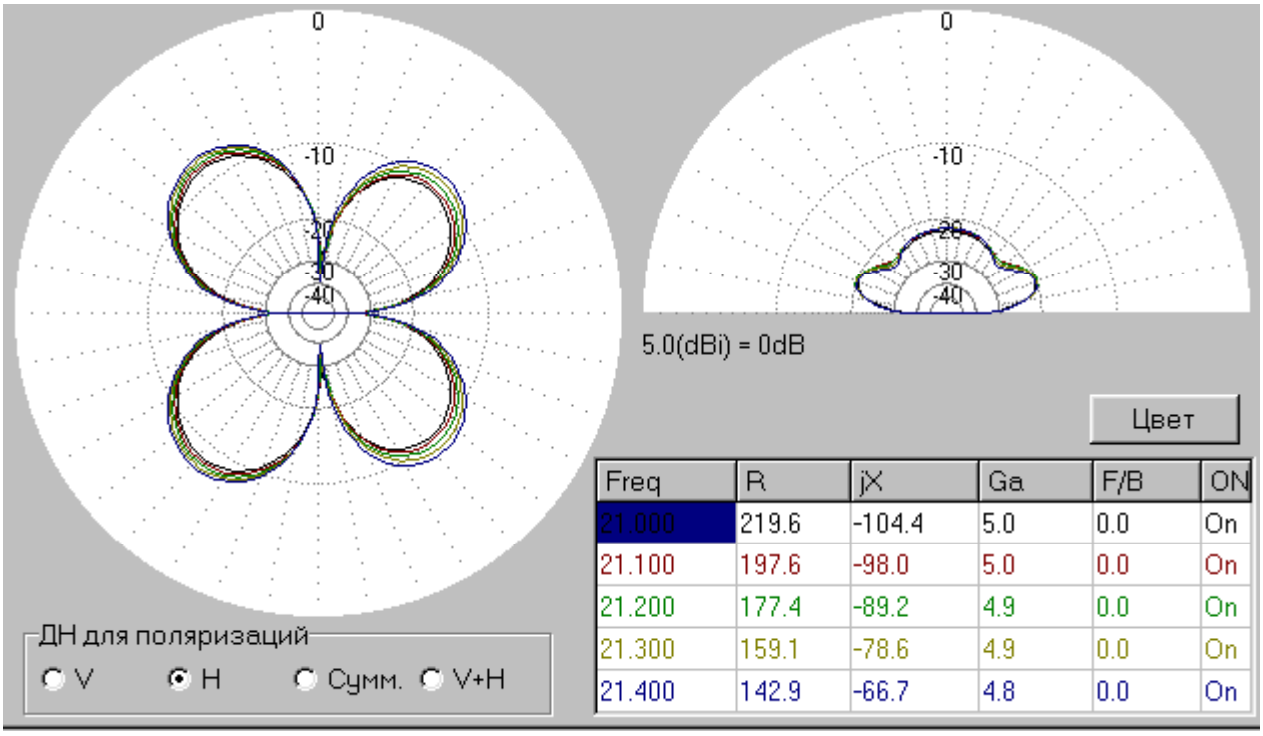
Shunt Vertical Universal HF Antenna at 15-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
21.000	219.6	-104.4	5.0	0.0	On
21.100	197.6	-98.0	5.0	0.0	On
21.200	177.4	-89.2	4.9	0.0	On
21.300	159.1	-78.6	4.9	0.0	On
21.400	142.9	-66.7	4.8	0.0	On

Horizon Radiation Pattern

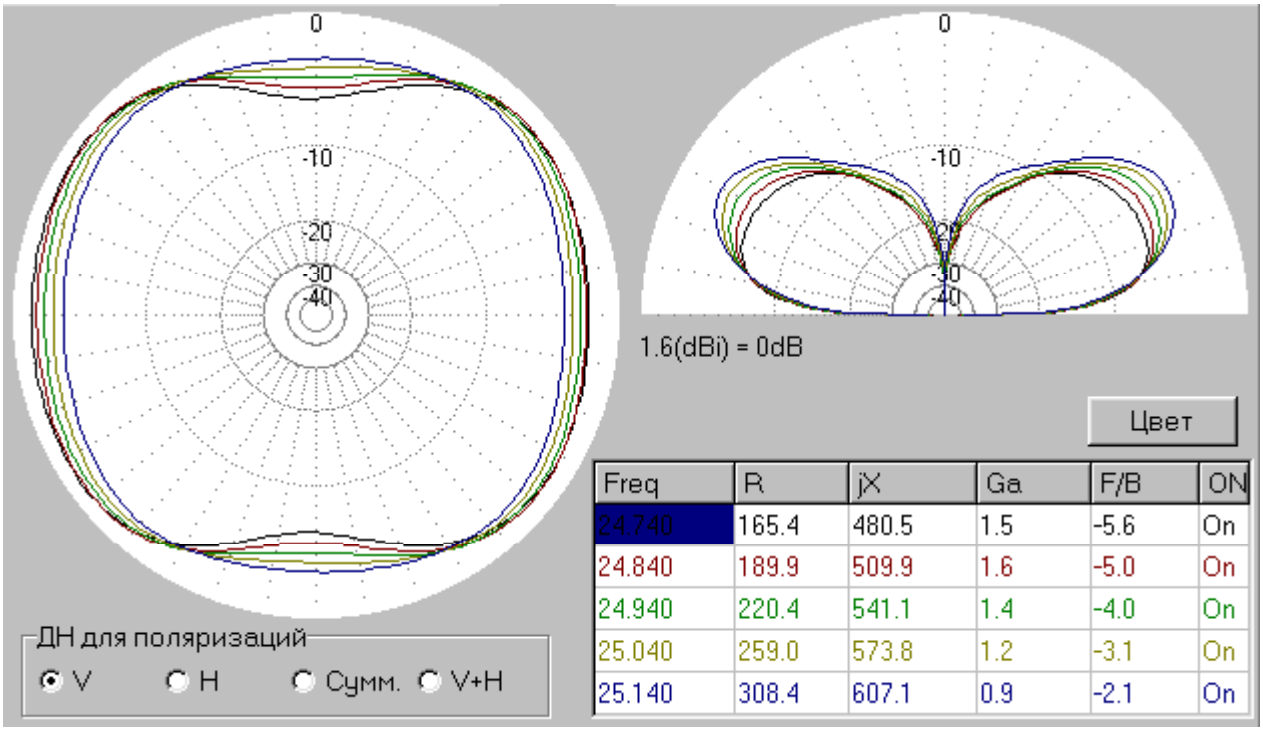


Freq	R	jX	Ga	F/B	ON
21.000	219.6	-104.4	5.0	0.0	On
21.100	197.6	-98.0	5.0	0.0	On
21.200	177.4	-89.2	4.9	0.0	On
21.300	159.1	-78.6	4.9	0.0	On
21.400	142.9	-66.7	4.8	0.0	On

Comments: Antenna has strong vertical radiation. A good pattern in the vertical plane. Antenna has almost egg shape pattern dropped to four lobes in horizontal plane, so, it demands to choose a proper direction before an installation of the antenna. Antenna provides DX QSOs. Antenna has Z= 177- j89-Ohms at 21200-kHz. Any ATU does good matching for such load.

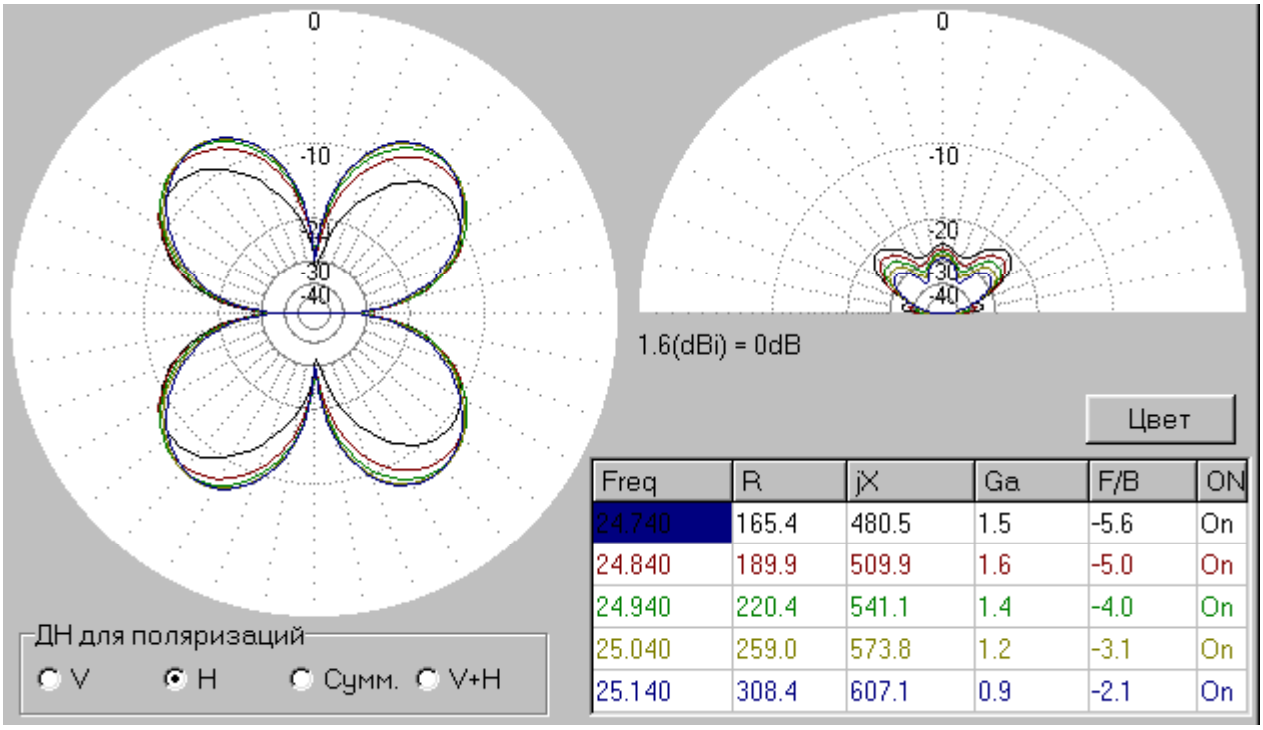
Shunt Vertical Universal HF Antenna at 12-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
24.740	165.4	480.5	1.5	-5.6	On
24.840	189.9	509.9	1.6	-5.0	On
24.940	220.4	541.1	1.4	-4.0	On
25.040	259.0	573.8	1.2	-3.1	On
25.140	308.4	607.1	0.9	-2.1	On

Horizon Radiation Pattern

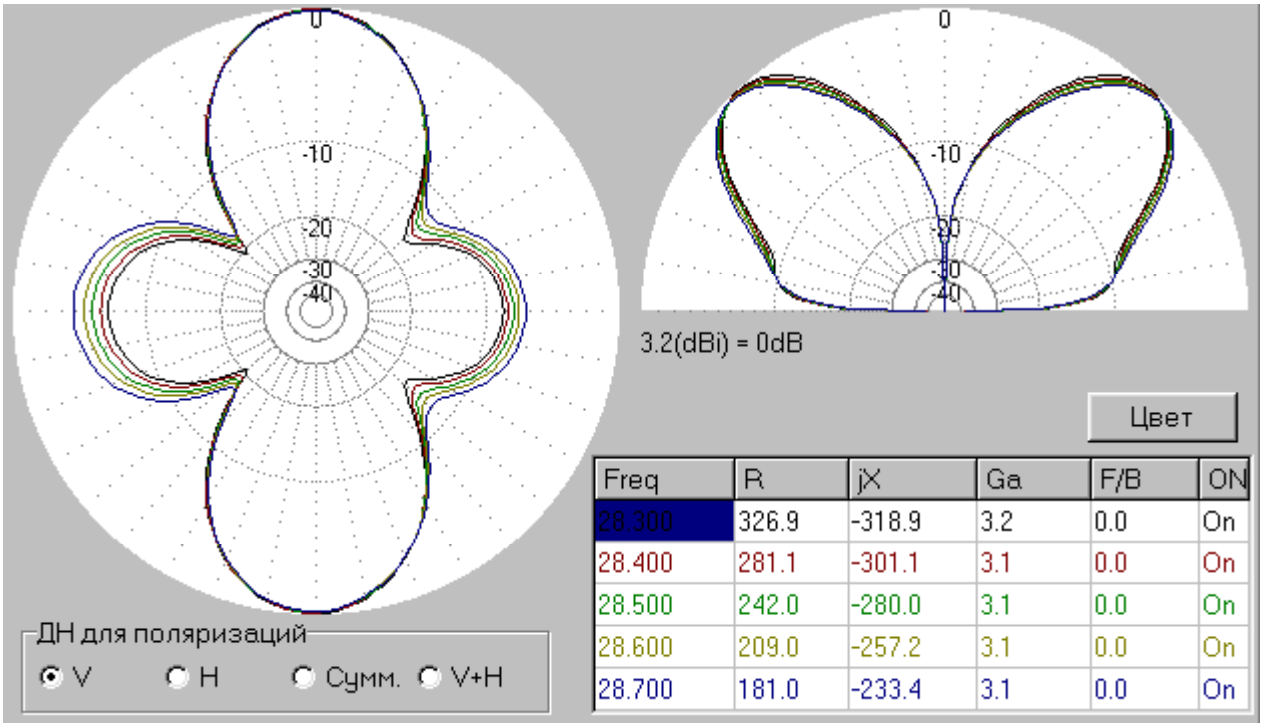


Freq	R	jX	Ga	F/B	ON
24.740	165.4	480.5	1.5	-5.6	On
24.840	189.9	509.9	1.6	-5.0	On
24.940	220.4	541.1	1.4	-4.0	On
25.040	259.0	573.8	1.2	-3.1	On
25.140	308.4	607.1	0.9	-2.1	On

Comments: Antenna has strong vertical radiation. A good pattern in the vertical plane. Antenna has almost circular pattern in horizontal plane. Antenna provides DX QSOs. Antenna has Z= 259- j573-Ohms at 25040-kHz. Not all ATUs do good matching for such load.

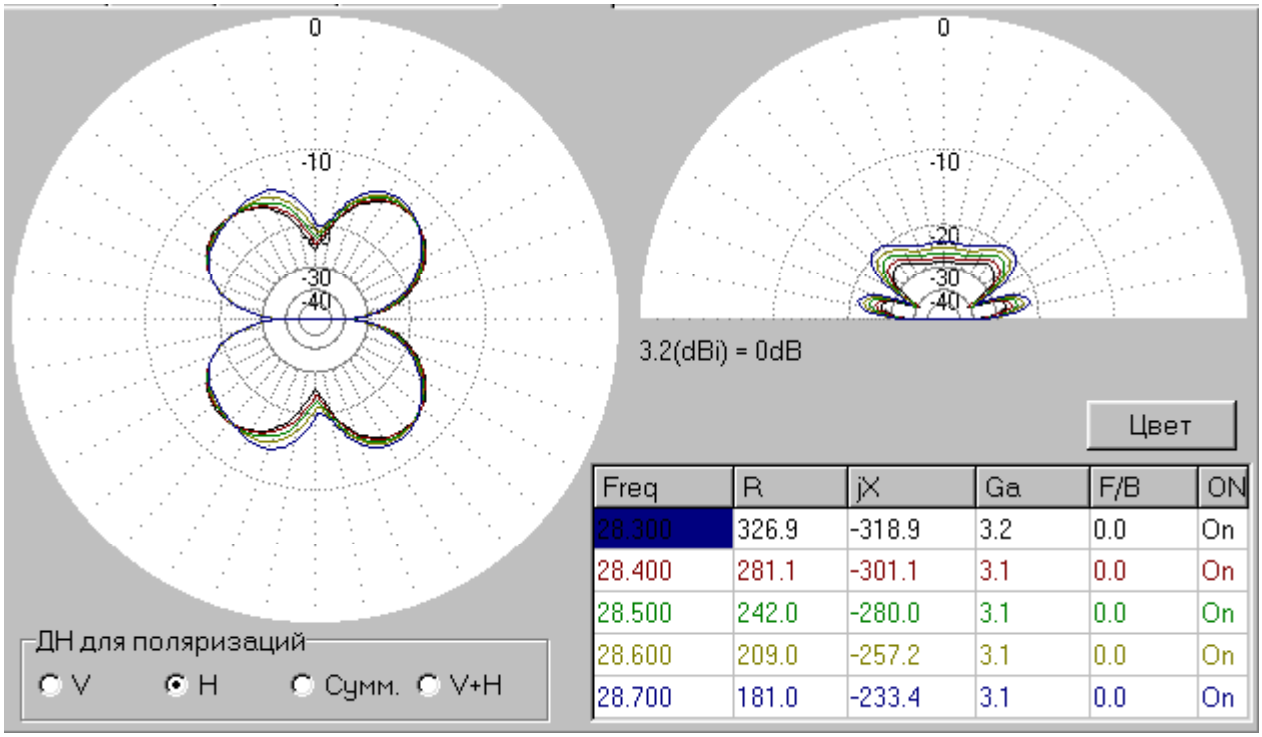
Shunt Vertical Universal HF Antenna at 10-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
28.300	326.9	-318.9	3.2	0.0	On
28.400	281.1	-301.1	3.1	0.0	On
28.500	242.0	-280.0	3.1	0.0	On
28.600	209.0	-257.2	3.1	0.0	On
28.700	181.0	-233.4	3.1	0.0	On

Horizon Radiation Pattern

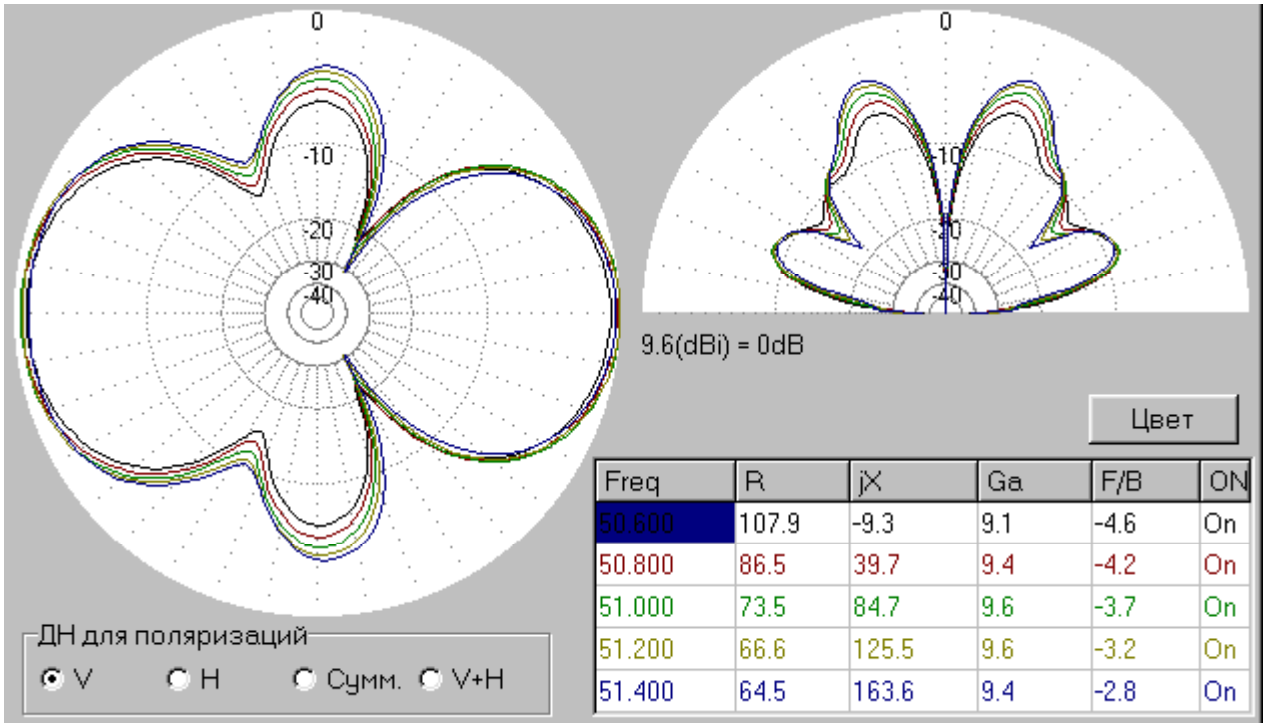


Freq	R	jX	Ga	F/B	ON
28.300	326.9	-318.9	3.2	0.0	On
28.400	281.1	-301.1	3.1	0.0	On
28.500	242.0	-280.0	3.1	0.0	On
28.600	209.0	-257.2	3.1	0.0	On
28.700	181.0	-233.4	3.1	0.0	On

Comments: Antenna has strong vertical radiation. A not bad pattern in the vertical plane. Antenna has a four- shaped lobes pattern in horizontal plane, so, it demands to choose a proper direction before an installation of the antenna.. Antenna can provide DX QSOs. Antenna has Z= 242- j280-Ohms at 28500-kHz. Any ATU does good matching for such load.

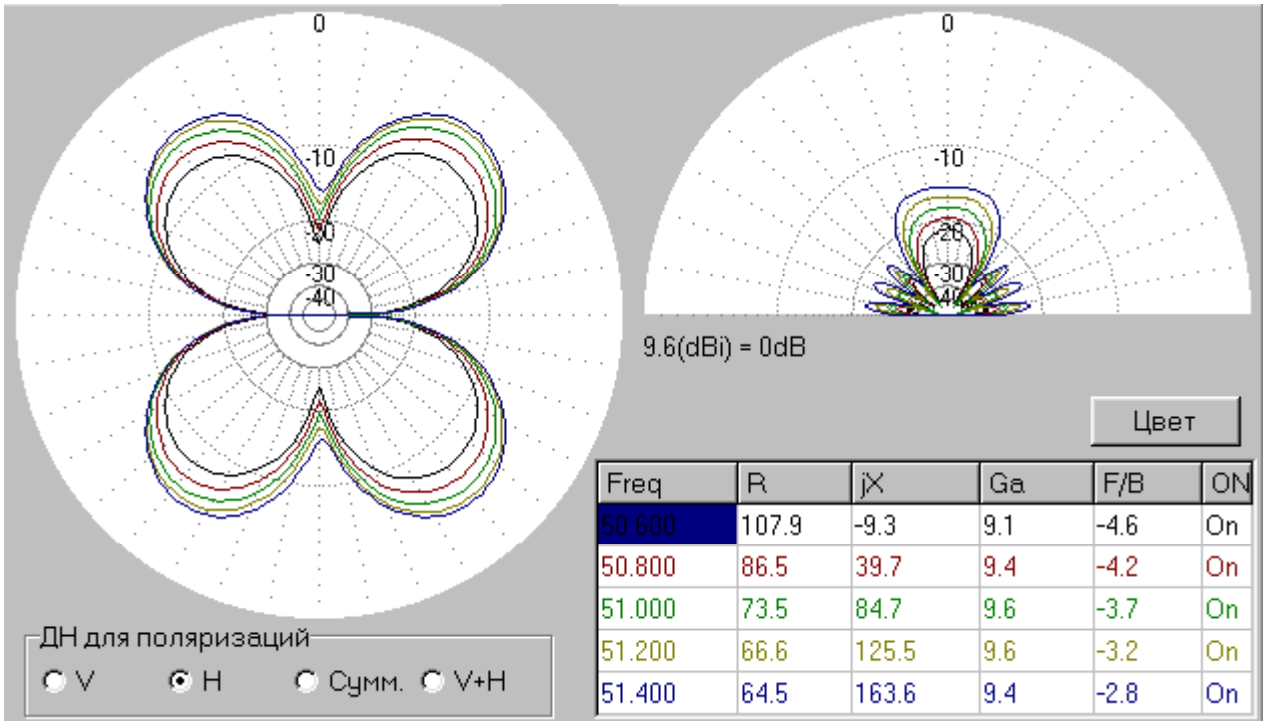
Shunt Vertical Universal HF Antenna at 6-m

Vertical Radiation Pattern



Freq	R	jX	Ga	F/B	ON
50.600	107.9	-9.3	9.1	-4.6	On
50.800	86.5	39.7	9.4	-4.2	On
51.000	73.5	84.7	9.6	-3.7	On
51.200	66.6	125.5	9.6	-3.2	On
51.400	64.5	163.6	9.4	-2.8	On

Horizon Radiation Pattern



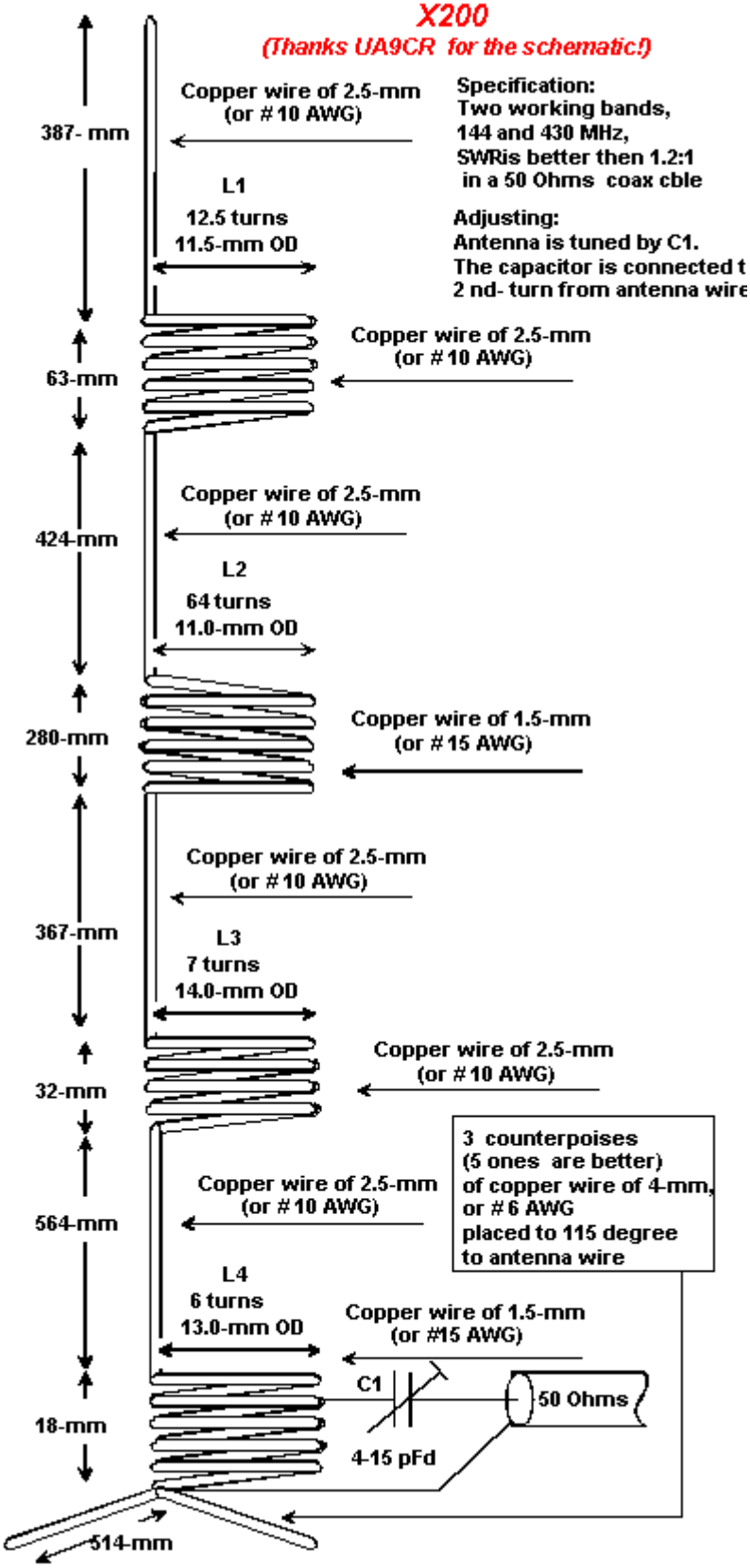
Freq	R	jX	Ga	F/B	ON
50.600	107.9	-9.3	9.1	-4.6	On
50.800	86.5	39.7	9.4	-4.2	On
51.000	73.5	84.7	9.6	-3.7	On
51.200	66.6	125.5	9.6	-3.2	On
51.400	64.5	163.6	9.4	-2.8	On

Comments: Antenna has strong vertical radiation. A not bad pattern in the vertical plane. Antenna has an “eight- shape” pattern in horizontal plane, so, it demands to choose a proper direction before an installation of the antenna.. Antenna can provide DX QSOs. Antenna has Z= 86+ j40-Ohms at 50800-kHz. The resonance is at the 6- meters band. Any ATU does good matching for such load.

CHAPTER 2

VHF ANTENNAS

Antenna X200



Александр UA9CR

Dear Friends,

Most of us are heard about VHF antenna X200. It is very interesting and very reliable two bands antenna.

RV9CX made some modifications for the antenna, so, the new RV9CX-X200 is more suitable for doing at amateur conditions. Go to the next page for the new antenna!

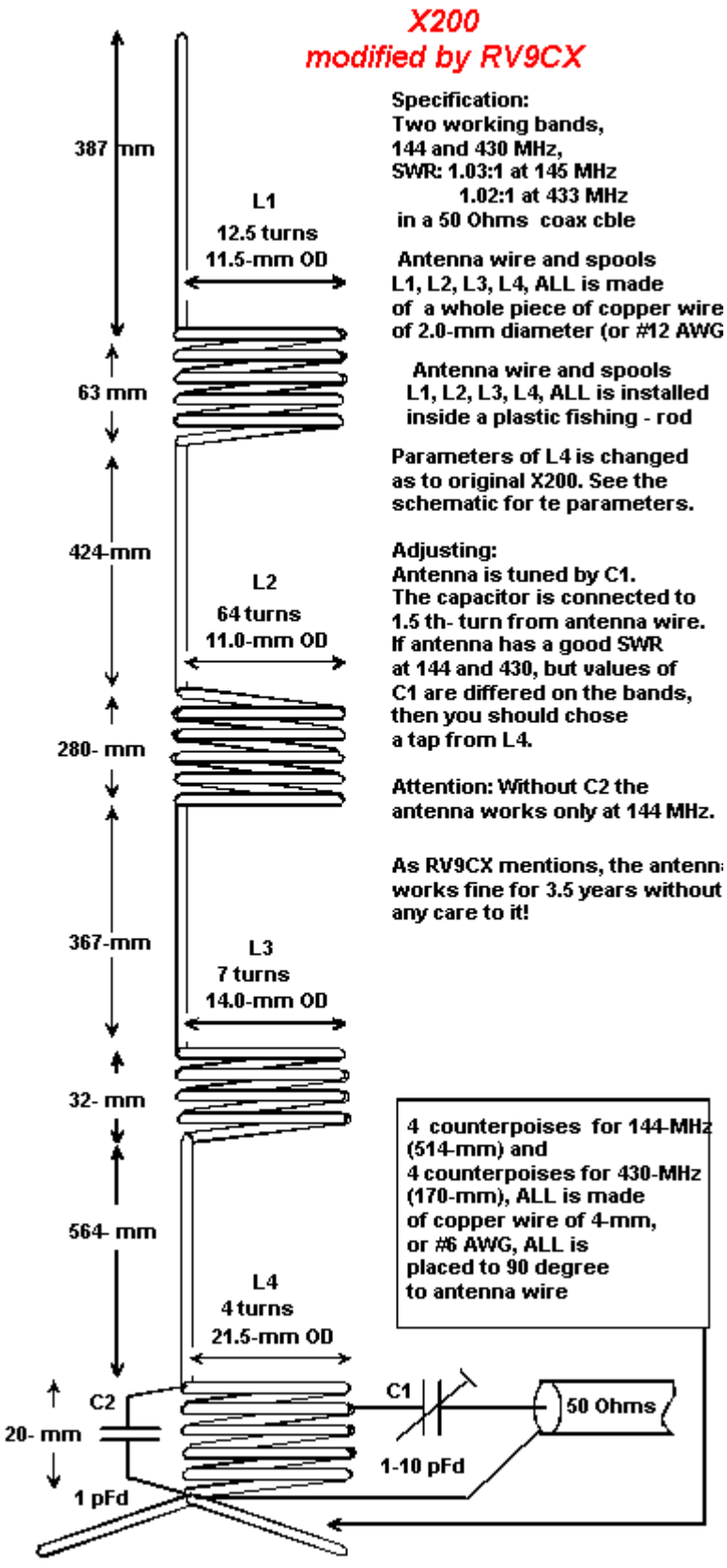
Comments about the antenna please send to Dmitriy, RV9CX: rscs@rosteck-msi.ru

Also, you can visit to

www.znuki.ru

The site, where are Dmitriy and his friends

Antenna X200



"Bottle" Antenna for 145 MHz

By Sergey Mironov, RA1TW

Any amateur can do the antenna during one hour. To do the antenna takes a half of hour and to tune the antenna also takes a half of hour. So, do not waste time and go to make the Bottle Antenna!

At first take a look at schematic of the Bottle Antenna (**Figure 1**).

Specification:

1. Dielectric plate, approximately of 80x250-mm.
2. Vibrator, (5/8)λ
3. Matching spool
4. Tinned plate, approximately of 25x35-mm.
5. Counterpoises, (1/4)λ
6. Stud, washers, screw-nuts
7. Coaxial cable

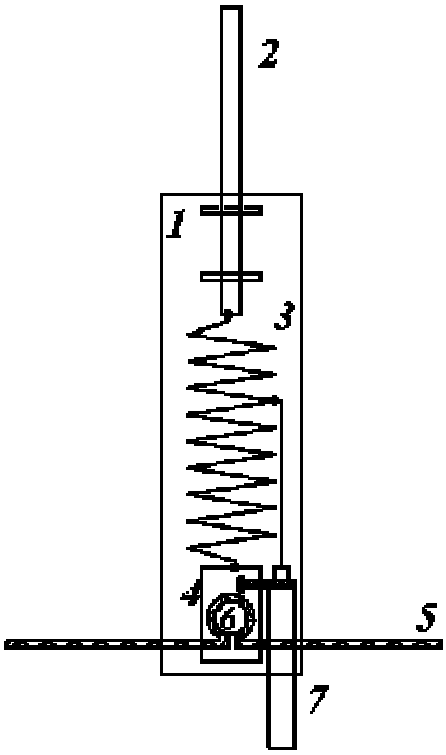
How to do it

1. Take Dielectric plate (1) and install Vibrator (2) on the plate. Use clamps or hard wire for this.

145 MHz the vibrator takes the length 1270-mm at the diameter of the vibrator 4…5-mm, and 1200 m at the diameter of the vibrator 10…14-mm.

2. Install Tinned plate (4) on the Dielectric plate. Use Stud, washers, screw - nuts (6).
3. Do Matching spool (3). The spool has 9 turns of 1.5…2.5-mm diameter (# 14- 10 AWG) copper or silvered plate wire. ID of the spool is 15…18-mm, RA1TW use to old markers as a form for the spools. Length of the spool is 34-mm.
4. Install the Matching spool on the Dielectric plate. For doing this, the upper end of the spool is fixed to the Vibrator and the down end of the spool is fixed to the Tinned plate. Use solder or fix the ends with the help of screws.
5. Do counterpoises. Two 105-cm lengths of copper or aluminum wire of 4…6-mm (# 2…6 AWG) are bended as a Greek letter OMEGA looks (see **Figure 2**).

Bottle Antenna
Side view



Top view

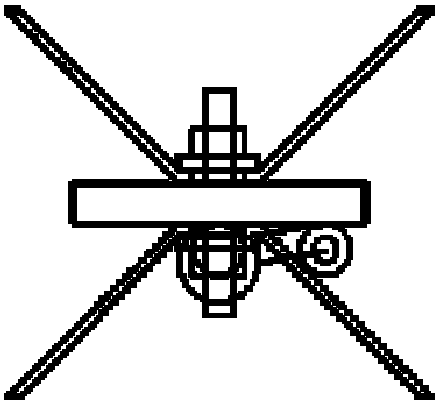


Figure 1

Figure 2. Counterpoises

6. Fix the OMEGA- counterpoises to the Stud with the help of the Screw-nuts. The counterpoises should be at 90 degree to the Vibrator and to each other.

7. Fix the OMEGA- counterpoises to the Stud with the help of the Screw-nuts. The counterpoises should be at 90 degree to the Vibrator and to each other.

8. Install Coax (7). Central core is soldered to 3-1/3 tap from the Vibrator, the braid is soldered to the Tinned plate (4).

9. Take a Bottle from dry drink. (I know, RA1TW always prefers a bottle from beer!. I.G.) Do a hole in the screw - top of the Bottle, cut the bottom of the Bottle, cut four slots for counterpoises, and then, install the Bottle (see photo) on the Antenna.

That is all the Bottle Antenna is ready!

Tuning:

The best way to adjust the Bottle Antenna is to use Meter of Amplitude vs Frequency Response characteristics. The device is switched to the Bottle Antenna and we see the frequency characteristic of the antenna. Stretch out the Matching spool or cut lengths of the Vibrator and Counterpoises if the resonance frequency of the antenna is below then 145 MHz. Gripe the Matching spool if the resonance frequency of the antenna is higher then 145 MHz Then select the tap for the best SWR. It is possible to match with the antenna a coaxial cable with any characteristic impedance –50 or 75 Ohms.

RA1TW

Of course, you can adjust the antenna with the help of only SWR – meter or VHF - bridge.There are 8 such home - made Bottle Antennas at Novgorod. All antennas work very well. They provide good communication as inside city as from city to a country for a distance more of 100 kilometers.

Credit line: http://hamnv.boom.ru

5/8 λ VHF- UHF ANTENNA

Alex, RA3GBQ, hamradio.lipetsk.ru, natan@lipetsk.ru

To do the antenna one can very easy as well as the costs are nothing. You need a box of a sweets- surprise Chupa- Chups, a piece of an old coax, some wire, epoxies, and a little of job.

So, go to do it!

►Take glass-reinforced plastic rod 5 and upper part of a sweets- surprise Chupa- Chups 10.
Insert the rod and wire 4 for matching coil into item 10.

►Take a metal tube, cut a ring 7 and insert into item 10.

►Take a plastic cap 8 from a plastic can, fix a RF-socket 9 on it, solder the wire of the coil 4 to the socket , solder by 3-5 wire the ring 7 to the socket.

►Turn up item 10, fill up it by epoxies 6, and close it by the cap 8.

►When the epoxies are hardened, drill holes for counterpoises 11 at low side of the item 10 and through out item 7.

►Cut a thread into item 5 and onto counterpoises 11.

►Remove braid 3 from a coaxial cable, put on the braid 3 onto item 5 that before is covered a glue. The length of the braid 3 is equal to 5/8λ for 144 0r 430 MHz.

►Coil the spool 4 (any number of turns) with step between turns of 1-mm, solder the spool to the braid 3.

►Insert 6 λ/4 counterpoises.

►Meter SWR. Chose a tap from the coil 3 to minimum SWR.

►Put on a can of a pen onto the upper end of the rod 5.

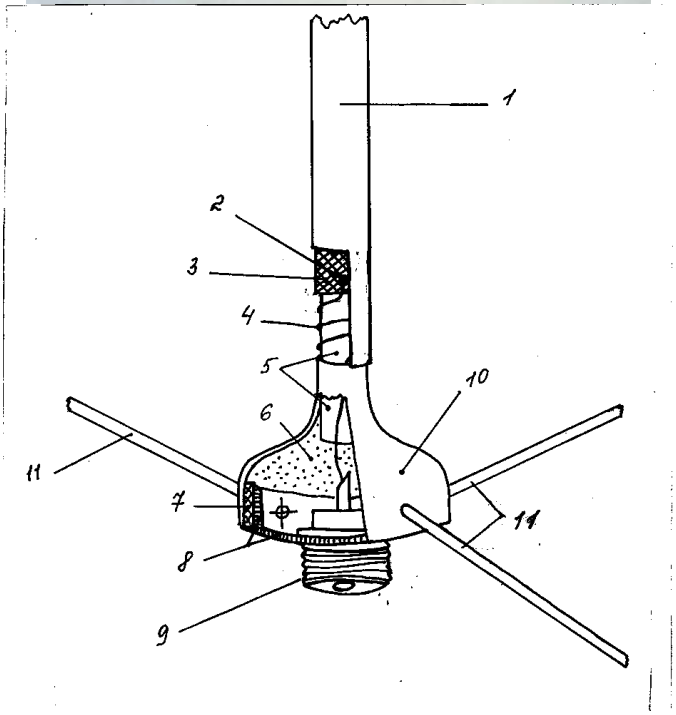
►Put on a shrink plastic tube 1 onto the rod 5.

►Heat the shrink plastic tube.

Get you own antenna!

73!

Sweets- surprise Chupa- Chups



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ANTENTOP

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SHORT 'RUBBER DUCK' FOR VHF/UHF HAND - HELD

Igor, UA6HJG, ua6hjg@mail.ru

All portable hand – held radios have a short "rubber duck" antenna. Some of the rubber duck antennas are rather long and if it is not necessary to communicate on a far distance, the long antenna just hinders to use a hand- held. I decided to make a short rubber duck antenna that is convenient for daily usage and for short distance communication.

Range 144 - 146 МГц

... For Yaesu VX-1R and C - 508

I have made this antenna for **VX-1R** as it has a small RF-socket SMA. Antenna has length in 43-mm (together with RF-socket). Copper wire of diameter 0.6-mm (# 23 AWG) is coiled up a turn- to turn on a form of diameter of 8-mm. 25.5 turns were coiled. At thus the resonance of the antenna was at 145.3-MHz with SWR 1:1. After winding and checks of the resonance, put on a shrink plastic tube on the antenna, and heat it.

For any stations having BNC socket

Antenna has length in 69-mm (together with RF-socket). Copper wire of diameter 1.3-mm (# 16 AWG) is coiled up a turn- to turn on a form of diameter of 10-mm. 22 turns were coiled. At thus the resonance of the antenna was at 145.3-MHz with SWR 1:1. After winding and checks of the resonance, put on a shrink plastic tube on the antenna, and heat it.

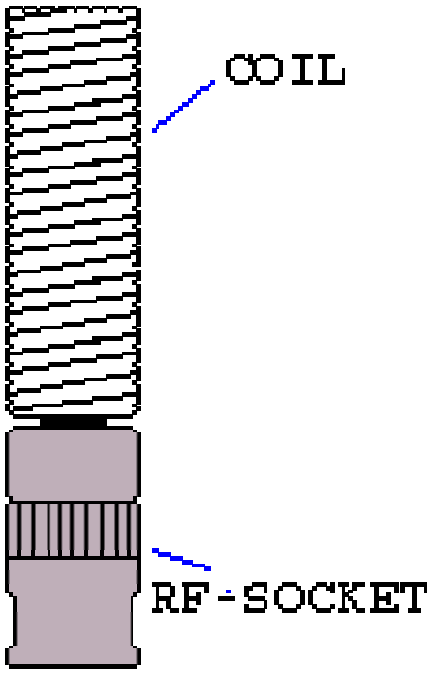
Range 430 - 440 МГц

The design of an antenna for a range of 433-MHz does not differ from the previous design, only it need to reduce the number of turns of the coil. Antenna has length in 42-mm (together with RF-socket). Copper wire of diameter 1.3-mm (# 16 AWG) is coiled up a turn- to turn on a form of diameter of 8.5-mm. 6 turns were coiled. At thus the resonance of the antenna was at 435.3-MHz with SWR 1.8:1. After winding and checks of the resonance, put on a shrink plastic tube on the antenna, and heat it.

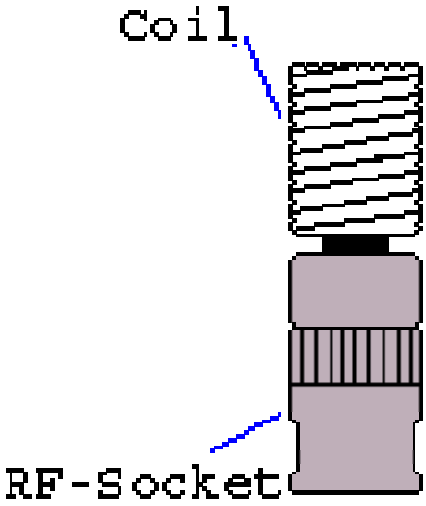
These designs were checked with hand-held radios VX-1R, C-568, FT-50 and C-508. At range of communication in city 1-2-kms the home- brew rubber duck almost do not lose to commercial – made ones.

73!

145 MHz 'Rubber Duck'



430 MHz 'Rubber Duck'



Credit Line: http://www.cqham.ru/

Twins Delta Loop for 145 MHz

*by **Nick V. Derenko**, **US8AR***

ex UB5AEO, UB4AR *mailto:us8ar@qsl.net*
http://www.qsl.net/us8ar
http://us8ar.narod.ru

UKRAINE

US8AR

SOMEWORDS ABOUTE MYSELF

Hello ! If it's interesting I'll tell some aboute myself. I was born 25 may 1957 y. in Snesznoe town of Donetsk obl. of Ukraine. After finishing Primorsko-Akhtarsk's school in 1974 I worked a locksmith and a turner. In 1975 I entered in Taganrog radiotechnical institute and finished it in 1980. After it I living and working in Romny town of Sumy oblast of Ukraine. I worked an engineer in Romny's branch of Leningrad NPO "Krasnaja Zaria" and seniorengineer in Special Design Bureau "Poisk". Since 1997 I working a foreman of powerenergetics of "Akhtyrkaneftegaz". I am married and have daughter and son.

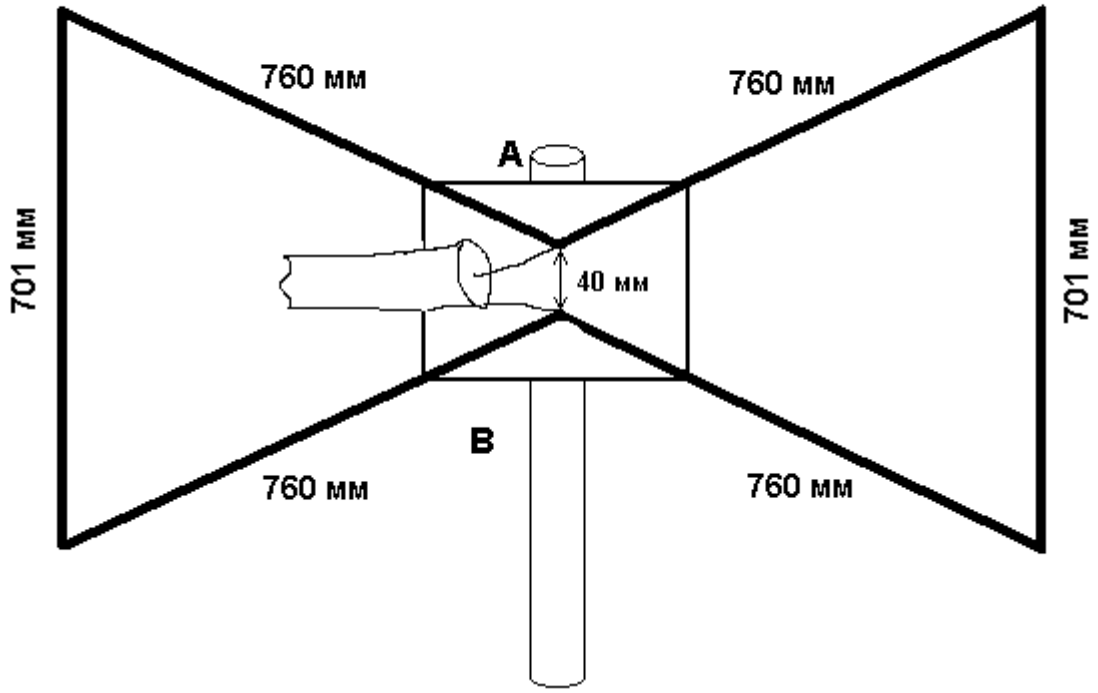
Specification:

- Directional diagram: "Eight" with low-altitude beam to horizon;

- Input resistance:	50 Ohm;
- Polarization -	Vertical;
- Gain	6 dB;
- SWR	1,01:1.

Figure 1 shows the antenna.

Figure 1



The antenna is made from a copper or aluminum wire in diameter of 4 mm. A copper or brass tube also will do well. The wires fastened by collars to a dielectric plate in 4 mm thickness. I use a plate from PC stuff. The plate fastened by collars to the antenna metal mast.

A coaxial cable is connected to points "A" and "B" (the central core to "A", the braid to "B").

Below, there is a file of the antenna in **MMANA**.

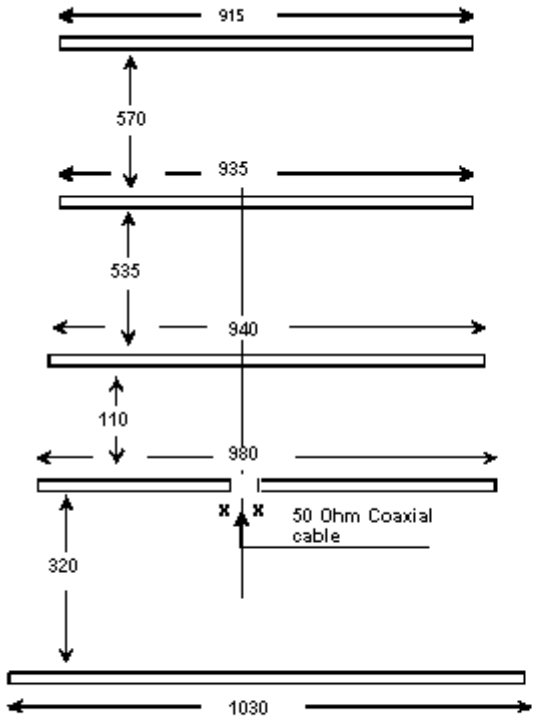
(MMANA available FREE at www.qsl.net)

The great collection Antenna Files at:
http://www.qsl.net/dl2kq/mmana

```
Twins Delta For  145_50 MHz
*
145.5
* wire *
7
0.0,      0.6845,  -0.3505, 0.0,      0.6845,  0.3505,  0.002,    -1
0.0,      0.6845,  -0.3505, 0.0,      0.0,     -0.02,   0.002,    -1
0.0,      0.6845,  0.3505,  0.0,      0.0,     0.02,    0.002,    -1
0.0,      0.0,     -0.02,   0.0,      -0.6845, -0.3505, 0.002,    -1
0.0,      0.0,     0.02,    0.0,      -0.6845, 0.3505,  0.002,    -1
0.0,      -0.6845, -0.3505, 0.0,      -0.6845, 0.3505,  0.002,    -1
0.0,      0.0,     -0.02,   0.0,      0.0,     0.02,    8.000e-04,      -1
*** sources ***
1,        1
w7c,      0.0,     1.0
*** load ***
0,        1
*** Autosegment ***
400,      40,      2.0,     1
*G/H/M/R/AzEl/X*
0,        5.0,     0,       50.0,    0,       0,       0
```

Enjoy!

Five Elements VHF Antenna RN1NZ For 145 MHz



Gain: 8.5 dBi

SWR: Less then 1.5:1 at 144...146 MHz

Front/Back Ratio: more then 18 dB

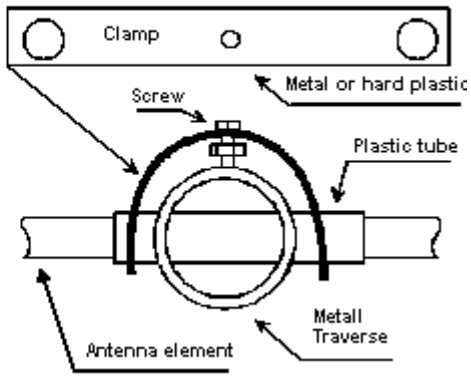
Input Impedance: 50 Ohm

All elements have diameter of 4-mm (or # 6 AWG)

Traverse has diameter of 15-mm and lenfth of 1600-m

Antenna elements are electrically insulated from the traverse

Antenna Construction



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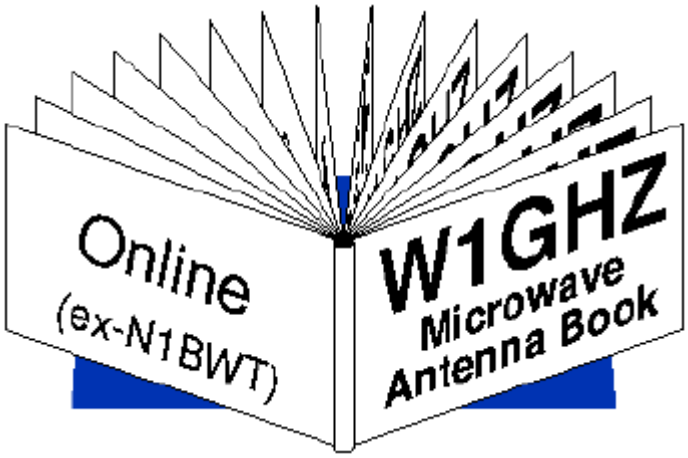

http://www.qsl.net/n1bwt/contents.htm

RN1NZ
rn1nz@onego.ru

Credit Line: RN1NZ @ Radio #4, 2002, p.65
For more info see:
www.radio.ru/

РАДИО

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Part-I: Practical Antennas (8 Chapters!)

Part-II: Antenna Measurement (2 Chapters!)

Part-III: Computer Analysis of Antennas (2 Chapters!)

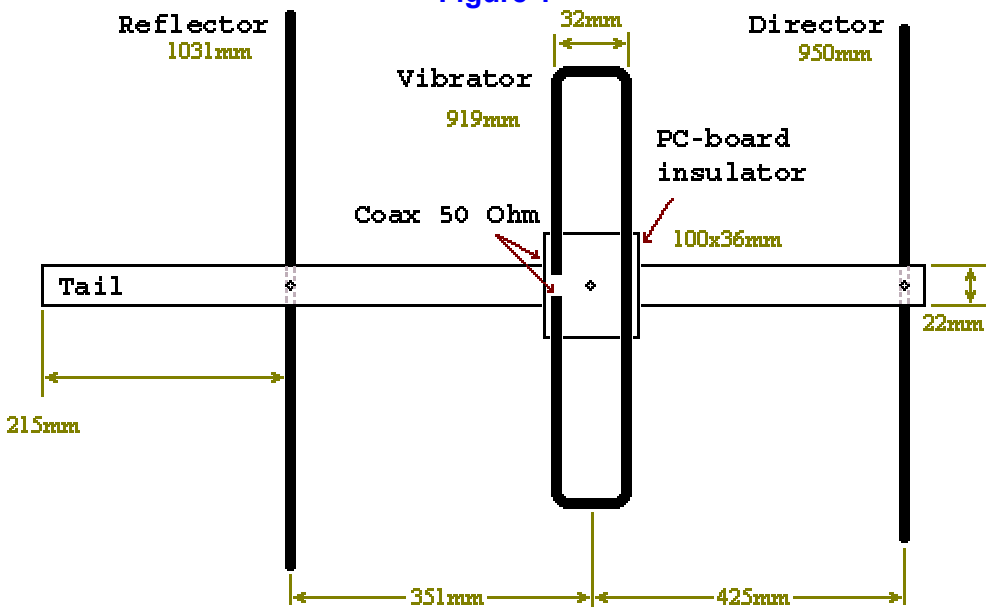
FOLDING 3-EL YAGI FOR MOUNTAINS

Igor, UA6HJG, ua6hjg@mail.ru

This antenna was developed specially for mountains trip from my experience of radio communication in mountains. The antenna is optimized to the maxima forward gain. At work you can hold the antenna by the "tail" and direct the antenna to your correspondent using vertical or horizontal polarization.

Figure 1 shows the design of the antenna. Antenna boom is made of a plastic water pipe of 1000-mm of length and of 22-mm OD. Antenna reflector and vibrator both are made of aluminum wire of 5-mm of diameter. Antenna vibrator is made of bimetallic wire of 5-mm of diameter. The vibrator is soldered to PC –board that is the central insulator for the vibrator.

Figure 1



A 50-Ohms coaxial cable is used with the antenna without any symmetrical devices. The cable goes along the boom and get out from the tail. For antenna folding you can loose screws, turn antenna vibrator along the boom, and again strength the screws. Antenna director and reflector is removed from the boom and hide inside the boom.

Coaxial cable has the length of 1100-mm. The cable is laid inside the boom and get out from the tail. It is need to protect from weather the place where the coaxial cable is soldered to the vibrator. RF socket also must be protected from weather and dirty at transportation of the antenna. Antenna has weight of 0.4-KG. Practical measured (by device SWR-121) SWR is: at 144.7=1.3:1, 145.2=1:1, 145.7=1.6:1.

Antenna has gain compare to half-wave dipole 6-7 dB. In general to measure the real gain of an antenna in the amateur conditions is practically not easy matter. I use a comparative method that shows advantage one antenna above other one. The essence of this method is simple. You take a dipole as the exemplary antenna and do compare the dipole with the experimental antenna.

Being on the southern slope of Elbrus- mountain at height of 4000 meters I use the method with help of Turkish ham station TA7T. The distance between me and TA7T was approximately 500-kms. I hardly heard TA7T by my exemplary dipole but I had 59 with the YAGI. It is near 10 dB in real gain!

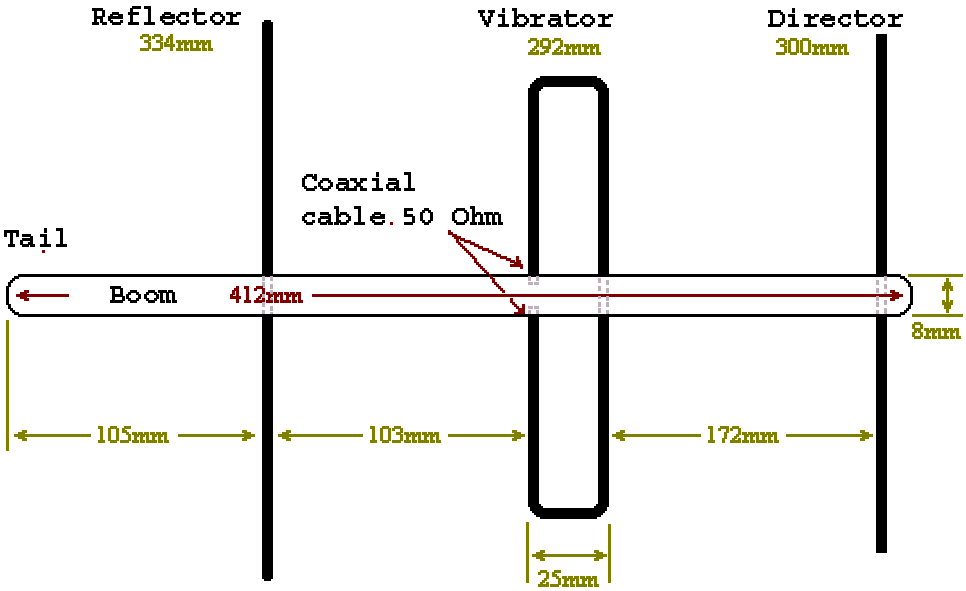
SIMPLE 430-MHz 3-EL YAGI FOR MOUNTAINS

Igor, UA6HJG, ua6hjg@mail.ru

I want to pay attention that UHF antennas for mountain climbing has the own specificity. So, antennas having the gain more the 6-9-Db-dB are not necessary in mountains at big height (from 3000 meters and is higher), as well as you do not need mast for the antennas. Also I prefer not knock- down antenna design. .

This antenna was developed specially for high-mountainous trip proceeding from my previous experience. At the first place I stand following characteristics: gain, weight, reliability. The antenna is optimized of the maxima forward gain. The design is made not knock- down that is

Figure 1



very conveniently as the antenna is always ready to operation. You can hold the antenna by its tail and direct the antenna to your correspondent at operation.

Figure 1 shows the design of the antenna. The boom is made of an ebonite rod of 412-mm length and of 8-mm diameter. Antenna elements are made of a copper wire of 2- mm (#12 AWG) diameter. 50 Ohms coaxial cable is soldered to antenna vibrator without any symmetrical and matching devices. The coaxial cable goes along the boom and get out from the antenna tail. Antenna elements are fastened to the boom by epoxies. The antenna can place above or inside a tourist backpack.

The coaxial cable has length of 50-mm. The place of the soldering of the coaxial cable to the vibrator must be protected from the weather. Antenna has weight of 150 gram. 150rp, real gain of 6-7 dB above half-wave dipole. SWR is: at 433-MHz =1.8:1, at 435-MHz =1.4:1, at 438-MHz=1:1, at 440-MHz=1.5:1 (it as measured by device 'SWR-121').

Commentary:
I recommend to use the antenna with low-power UHF radio (10-mWtts) for a range 433-434Мгц. I have 30- 50 kms of distance (at direct vision) using the antenna with such station. Also I recommend to use the antenna with usual UHF radios, be sure, you easy will skip more than 100-kms!

Credit Line: http://www.mountain.ru

SIMPLE 430-MHz 4-EL YAGI FOR MOUNTAINS

Igor, UA6HJG, ua6hjg@mail.ru

The purpose of this design is to create an antenna for mountains. The antenna must correspond follow requirements:

1. To have gain not less the 7-dB. Forward Gain is the main characteristic of the antenna.
2. A wide lobe must be. During a QSO you can hold the antenna by the tail and shaking of the antenna should not result to full 'failure' of the communication.
3. Simplicity in making and adjustments. An opportunity of fast repair in field conditions.
4. Antenna weight with coaxial cable both must be up to 500 gram.
5. 50-Ohm coax for the feeding must be used.

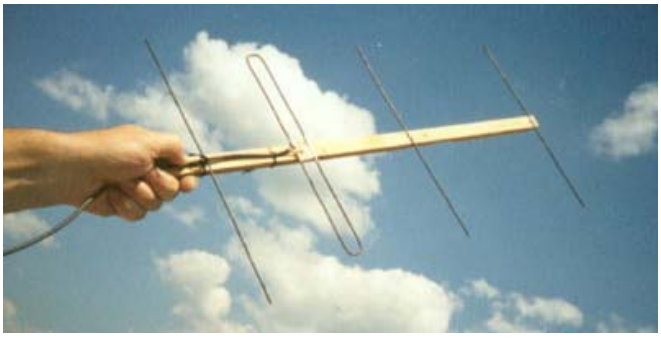
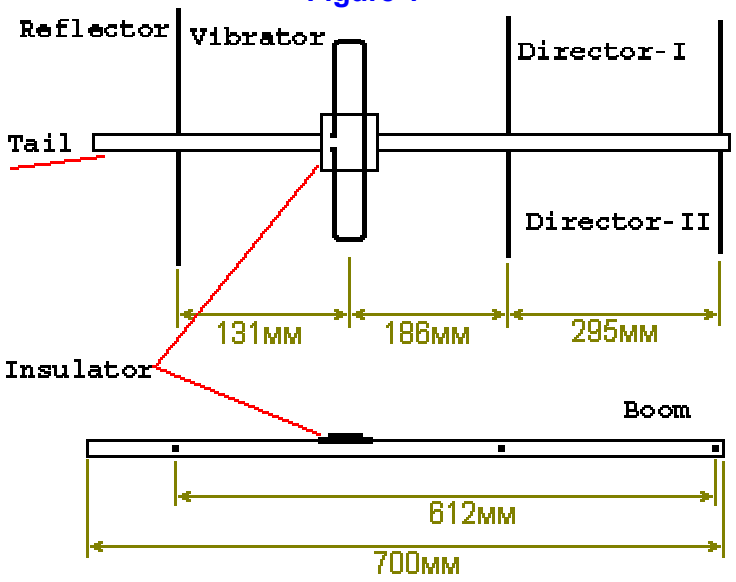


Figure 1 shows a design of such antenna. The boom is made of a wooden strip.. Elements and the vibrator are made of 2-mm copper wire (# 12 AWG).

The sizes of elements:
Reflector 345-mm.
Vibrator 285-mm.
Director-I 312-mm.
Director-II 306-mm.

Figure 1

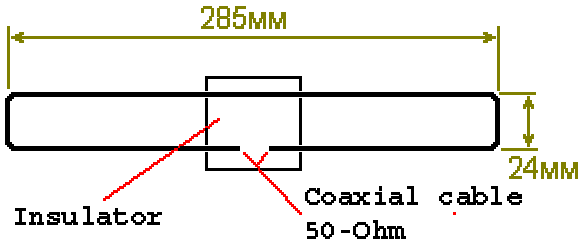


The design of antenna vibrator is shown on **Figure 2**. Coaxial cable has 850-mm length and this one goes along the boom. Soldered place must be protected from weather.

Antenna has weight of 290 gram. SWR (metered by SWR-291) is: at 433- MHz =2:1, at 435-MHz =1:1, at 438-MHz =2:1.

The antenna worked very well in the mountains. Also the antenna may be used for stationary work from a fixed QTH.

Figure 2

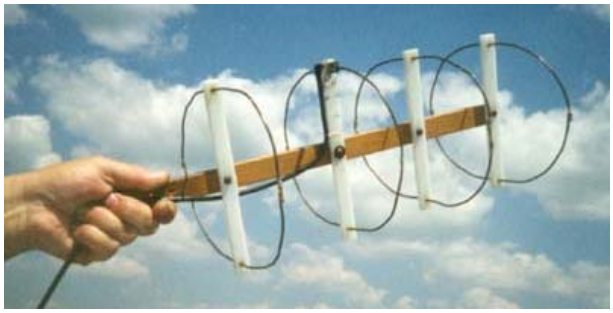


Credit Line: http://www.mountain.ru

4-Ovals Antenna for 430-440 MHz

Igor, UA6HJG, ua6hjg@mail.ru

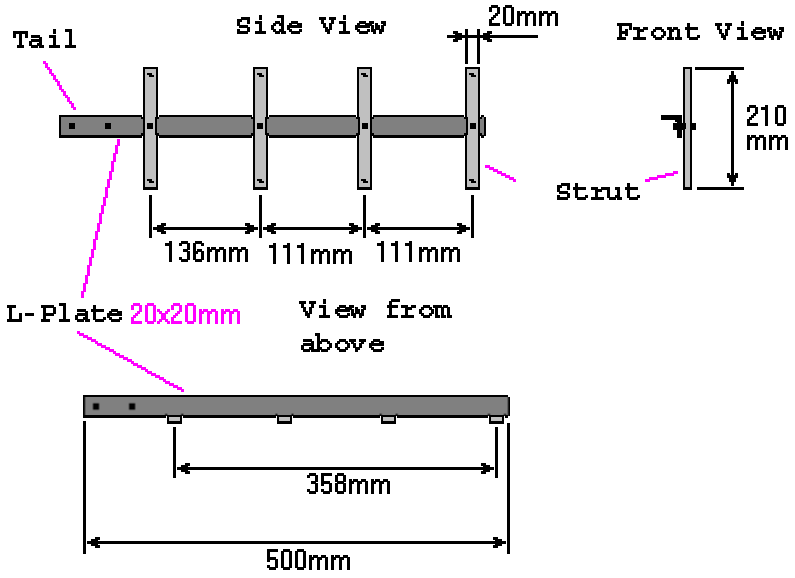
I need a good universal antenna for repeaters working at the 70-cm range, packet radio and for routine work. Such antenna must be satisfy the follow characteristics:



1. Real Gain 9-10-dB.
2. Low SWR at 430-440- MHz.
3. Wide forward lobel.
4. Unpretentiousness to close located subjects.
5. Simplicity in manufacturing and adjustments.
6. Possibility to create phased антенные systems on the basis of the antennas.
7. 50 - Ohm coaxial cable feeding.
8. Stability to heavy icing and to winds in 10-20 meters per second..

4 Ovals Antenna responded all of the conditions. Why an oval? At first, from the antenna theory we know that an oval radiates energy a little bit more effectively than a square. At the second, in practical, it is more easy to do an oval then a square.. So choose the OVAL! **Figure 1** shows a design of the

Figure 1



antenna. Boom is made of iron L-plate 25x25-mm. All the four ovals are fastened to the boom by struts made of insulation stuff such as tree, hetinax, etc. Antenna is fastened by the tail, that is behind the struts.

Perimeter of elements (ovals):

Reflector- 727-mm. *(-20-mm for soldering)*.

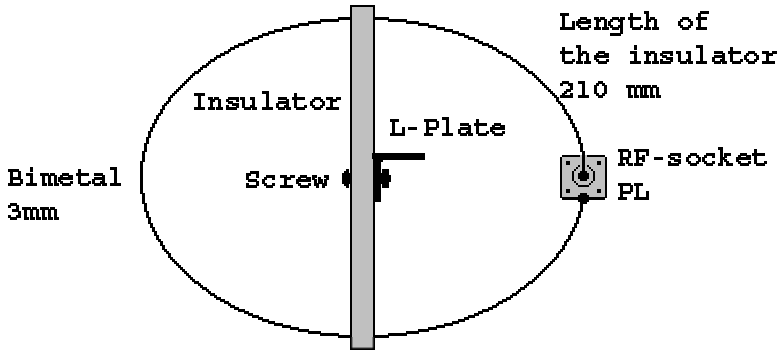
Driven Element- 653-mm *(+10-mm distance for a RF socket)*.

Director-I- 649-mm. *(-20мм)*.

Director- II - 639-mm. *(-20-mm)*.

Figure 2 shows the design of the Driven Element. RF socket is soldered directly by the ends of the oval. Three holes is drilled at each strut. One holes is in the center for fastened the strut to the boom, two holes are at the ends for bimetal oval. Antenna reflectors and directors have the same design as the Driven Element only difference is its ends soldered together.

Figure 2



Antenna has weight of 900 gram. SWR is: at 430-MHz =1.2:1, at 435-MHz =1:1, at 440-MHz=1.3:1 SWR was metered by device SWR-121. If you will use boom made of iron L-plate having not the same dimensions you would do correction lengths of antenna elements. I want to pay attention, that the antenna is more laborious than YAGI, but it work much better.

4-Ovals Antenna for 430-440 MHz for Mountains

Igor, UA6HJG, ua6hjg@mail.ru

When an antenna is intended for mountains it must follow such additional requirements as:

1. Light weight.
2. To have folding design and to take a little place in a backpack.
3. Antenna should be conveniently to hold in a hand, to fasten to a tree or stick. .
4. Tolerance to a dirty, water and snow.
5. Possibility to repair in field conditions.
6.

So, the boom of the trip antenna is made of a wooden strip drying by oil. Perimeter of all ovals is increased on 20-mm compare to previous design. All struts are made from un- foiled PC-board. Antenna elements are made of 2-mm diameter soft copper wire (# 12 AWG). Soldered parts must be protected from the weather.

Antenna has weight 400 gram. Other characteristics are as the previous design has. The antenna is tested at 1997 and at 2000 in the mountain Elbrus and the antenna shows good result. I made QSOs in distance of 200-300 kms with 59 for both ends!

Igor, UA6HJG, on Elbrus

Elbrus

FM Crystal Radios?

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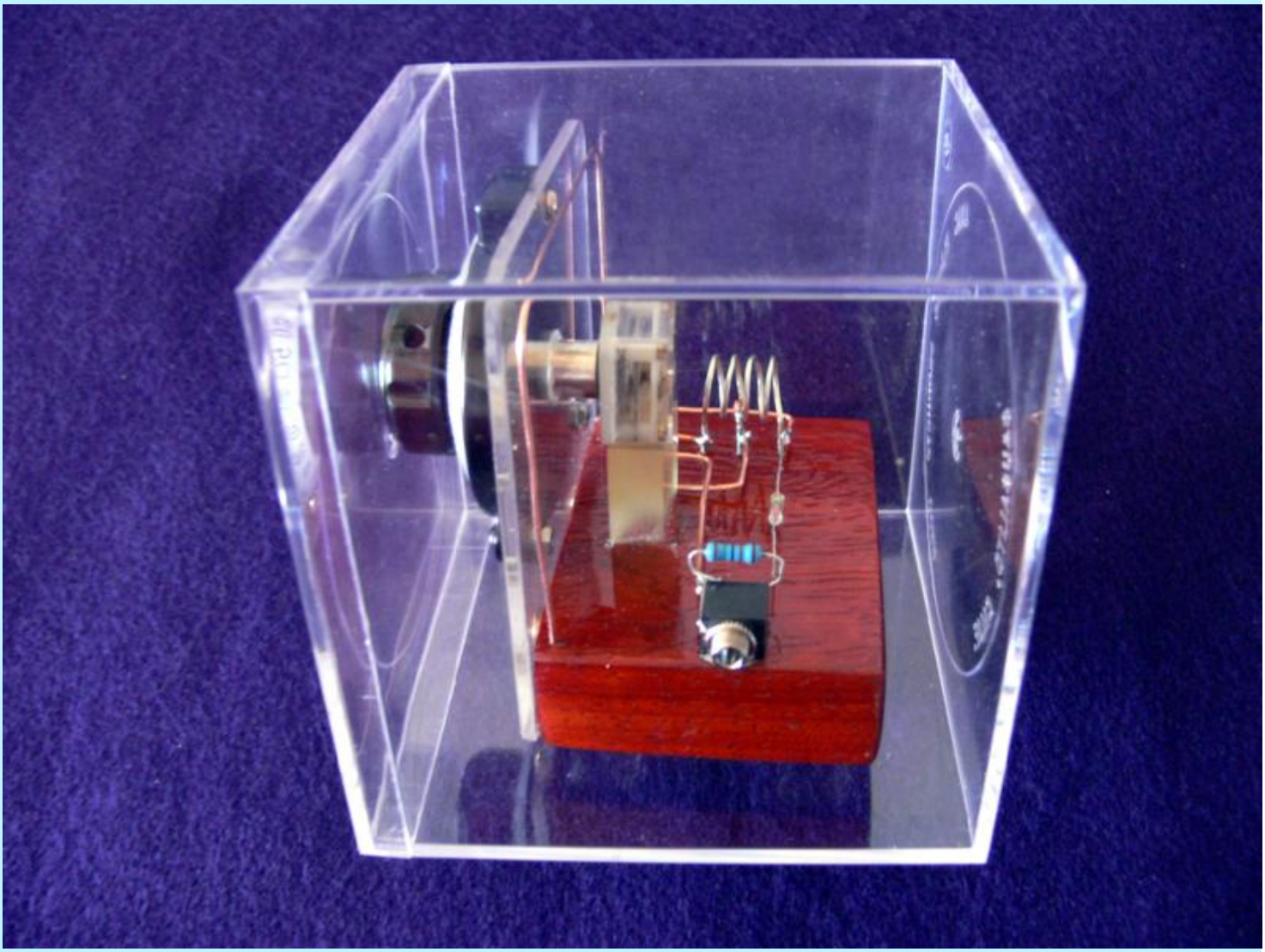
I have heard, even from a physicist, that it is impossible to build FM crystal radios. On the other hand some experimenters claim that they have built them. This argument intrigued me to try and build an FM crystal radio, which I have done successfully. To my surprise, the result is an astounding performer, pulling in four local stations in Tucson. When connected as a receiver to a good sound system the sound fidelity is as good or better than more expensive AM radios. In fact, it sounds "high-fidelity".



This picture shows the Solomon FM Crystal Set in an acrylic display case. I made the set specifically to fit inside this case (the case came first).

My definition of a crystal radio is one that is not powered, except by the radio transmission itself and employs a crystal detector. So, it should work without any batteries or AC power. An FM crystal receiver must be able to detect and receive FM signals well enough to be heard in earphones without any such extra power.

This FM receiver is an amazing performer. It has crystal clear reception (pun intended), good sensitivity, but only fair selectivity. This set was a discovery for me. I started out by designing and building the normal AM sets. Then one day while testing the "Mystery" set (see my other web links), to my surprise, in addition to the expected panoply of AM stations, I heard a very faint signal that I could not tune out. At first, it seemed too weak to identify. When I tuned out all the AM stations, I was astonished to hear the announcement "KiiM FM, 99.5"! This is a

country music FM station here in Tucson. It was all over the dial, untunable, but the much louder AM signals masked it when they were tuned in.

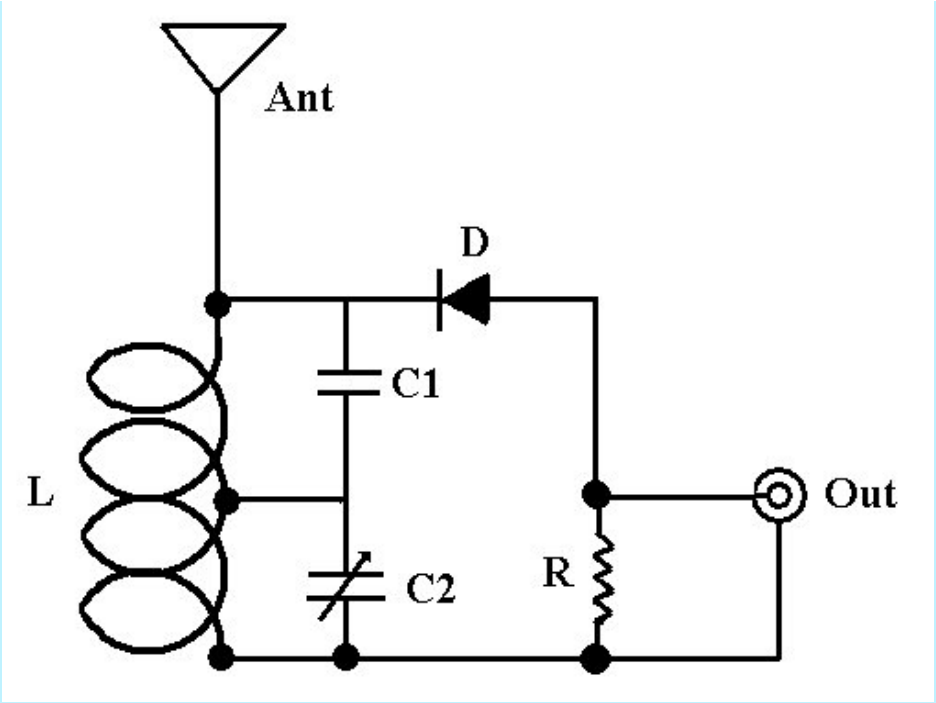
I set myself the task of trying to improve the FM reception. I tried some simple circuit modifications that did not seem to improve anything. Then I connected a dipole antenna instead of the AM antenna I normally use. Suddenly, the FM signal was much clearer, although still weak. By using the audio output and sound system amplifier, I was even more amazed that four different FM stations came in loud (or rather medium) and clear. I found that changing the telescoping antenna length and position I could tune the stations in and out. They were KRQ, KLPX, KiiM, and KHYT all local FM stations with transmitters nearby. Their reception was also affected by the length and position of the audio output cable.

After doing some research, I discovered that there was a physical theory that claimed that FM reception was possible and even probable using the same circuit as an AM receiver. The theory is called "slope detection". So, I set out to find circuit improvements. A web search yielded little, mostly theory. But there was enough information that I thought I could make some modifications to the AM circuits to make them more tunable to FM signals and less tunable to AM. Since FM operates at higher frequencies, all I had to do, I thought, was make the coil and caps smaller. After much "tinkering" I arrived at the current circuit.

The circuit looks identical to a classic AM crystal circuit but is even simpler to build. The components were reduced in dimension to resonate at higher frequencies. This was done by experimenting with smaller and smaller coils and capacitors. The antenna is also much reduced in size (from that of AM) to resonate at higher frequencies (the antenna is crucial). The air variable capacitor I used has two trimmers in it which should be adjusted for best reception.I have found that a commonly available vernier dial and knob will fit the capacitor nicely. See end of article for a picture of the variable. C3 is a ceramic capacitor of 18 pf, but may be anywhere from 10 to 50pf. A detected FM signal is converted to AM due to an effect called slope detection that modulates amplitude.

This FM Crystal Set works best near the transmitter (I have not tested it beyond about 10 miles). Secondly, the sound level is quiet, especially without an amplifier. A quiet room is needed for listening with earphones. One must be willing to move the set around to find a location for the best reception of signals. However, in addition to listening with high impedance earphones (crystal or otherwise), the set can be connected directly to an audio amplifier's low level magnetic input which can then play amplified through a sound system at any volume -- sounds GREAT. In fact, I recommend starting tests with the FM crystal set by connecting it to the low-level phono inputs of a receiver or preamplifier. (Nowadays, many receivers don't even have a phono input!) That way you can crank up the volume, which makes it more likely to find the FM stations. If no signals are detected, I also recommend connecting an external "rabbit ear" antenna or hanging a short wire (12 inches or so) in various positions next to the internal antenna. The variable length of rabbitt ears can help to tune in stations.

No additional wiring or antenna is necessary (the antenna is optimized in length for FM.)

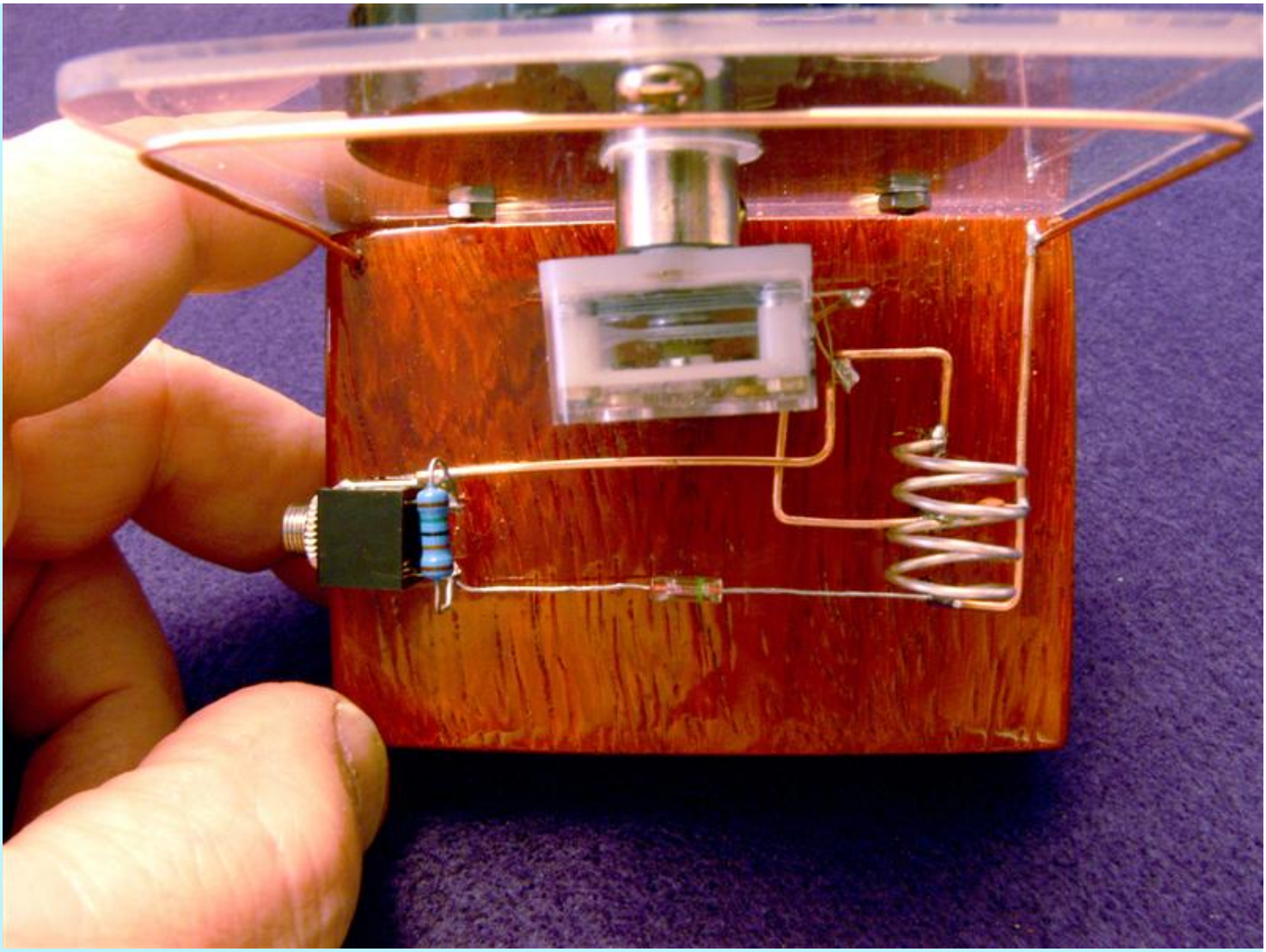


L - 4 turns #18 copper or silver wire, 12mm inside diameter, tapped at 2.5 turns
Ant - 7 inches of #18 bare copper wire
C1 - 18 pf ceramic capacitor
C2 - 50 pf air variable capacitor
D - 1N34 diode or rock crystal
R - 150K resistor

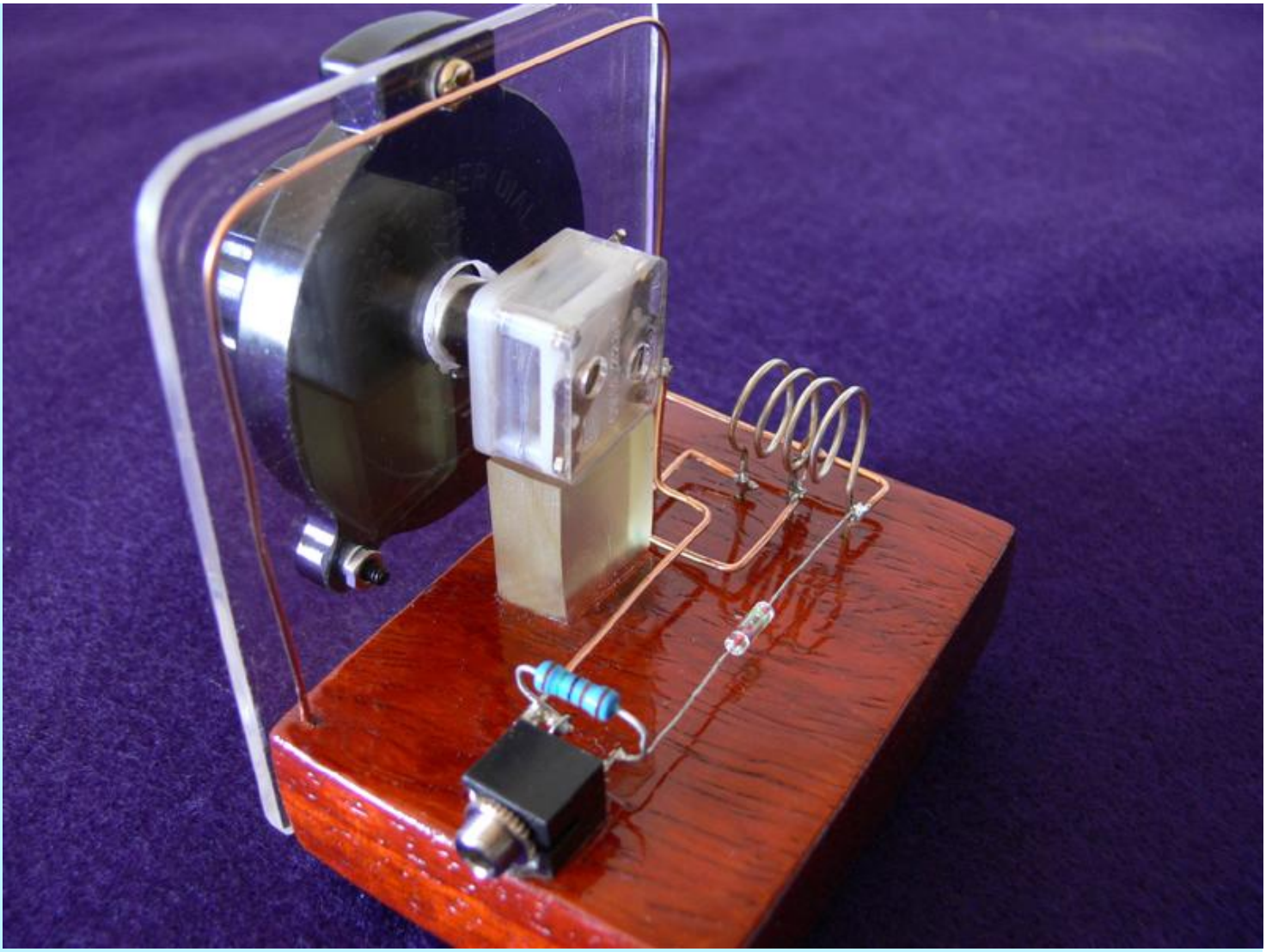
The diode is tapped directly to the antenna. The vernier dial fits directly on the tuning capacitor. The antenna parallels the perimeter of the acrylic face plate. "Military style" #18 AWG wiring is used without any insulation. It is important to keep the components physically close together. The component specifications are the same as in circuit #2. The coil is silver rather than copper, but copper does just as well. I think that the contrast of the silver and copper is beautiful. The coil was wrapped around a Sharpie Permanent Marker, then slipped off and expanded slightly. The wooden base is made from lacquered, polyurethane padouk.

I consider this set a work of art as well as science and think it is the most elegant crystal receiver I have created. I love the contrast of the silver coil, the copper antenna, the clear acrylic faceplate, the black vernier dial, the white and transparent variable capacitor, and the subtle colorings on the resistor, the diode, and the lucite base. Yet the circuit is so ridiculously simple that some will not believe it is possible without building it themselves. No shielding is necessary, and there is no problem with hand capacitance. However, the output cable position may affect reception sensitivity.

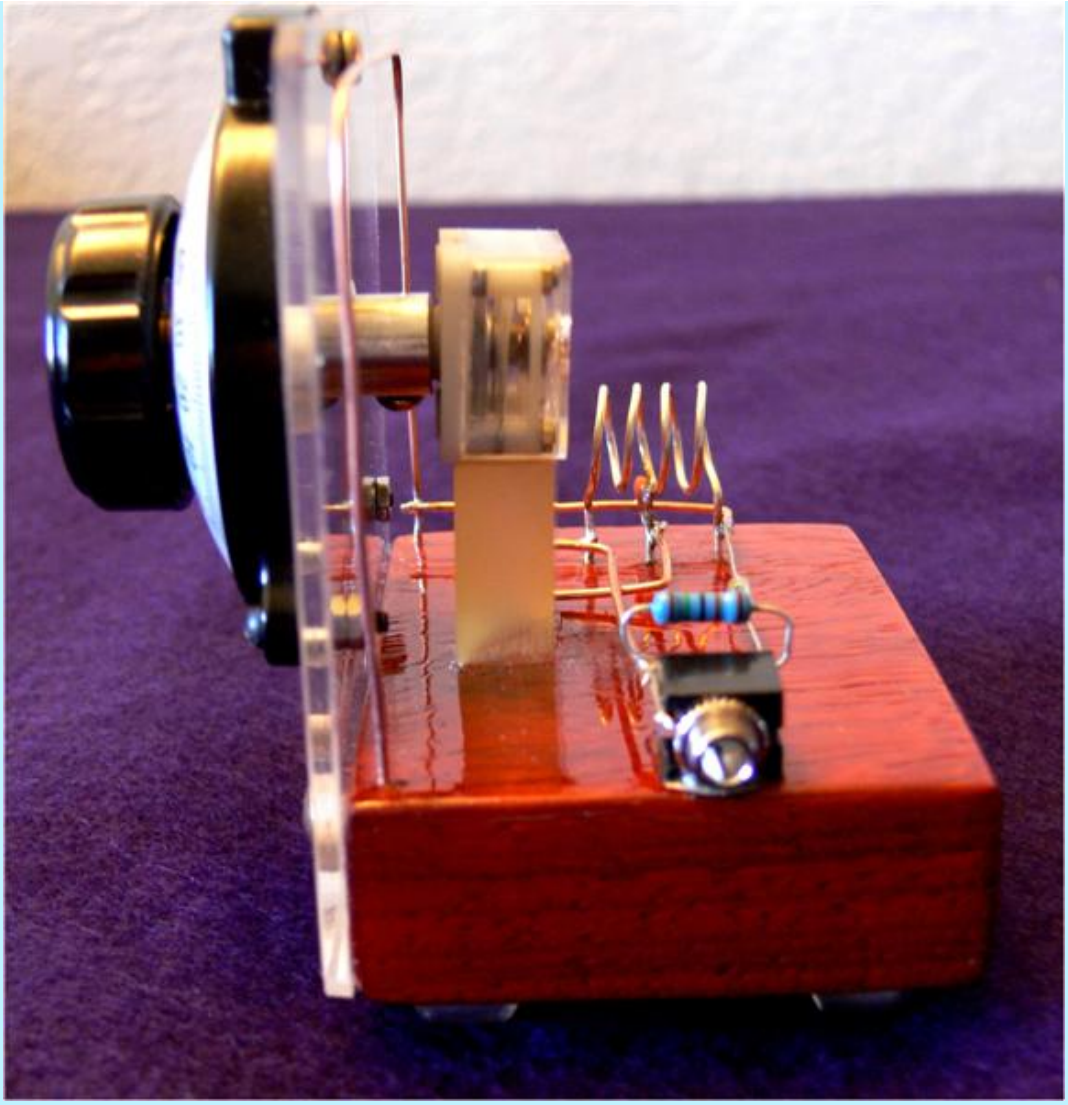
Photos of wired circuit



A hand is included in this photograph to show scale. Note the military style wiring, diode, and antenna. I wanted the wiring to create a modern design similar to a Mondrian painting. Not only is this set beautiful, it works! No power and no long antenna! It looks like a work of fiction.



Is this thing imaginary -- science fiction? Well, imagination did play a part, but it is definitely not science fiction. This shot shows the elegance of the FM set best, I think. There is only one resistor and one fixed capacitor.

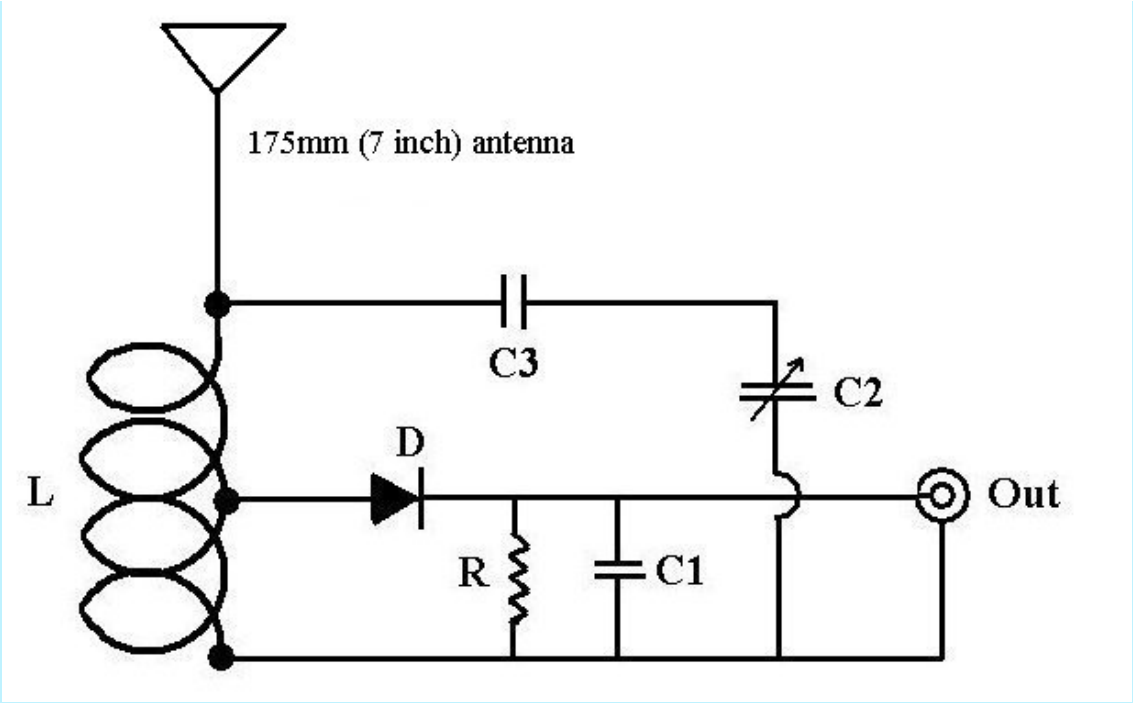


The inside of the tuning capacitor and the phono jack/output can be seen here. Can you spot the fixed ceramic capacitor? Note the polished edge of the face plate and the reflection in the wooden base.

A quarter-inch piece of lucite was fitted under the tuning capacitor to anchor it. Note the two tiny trimmers on the back of the tuning capacitor. Brass screws were used to enhance appearance.

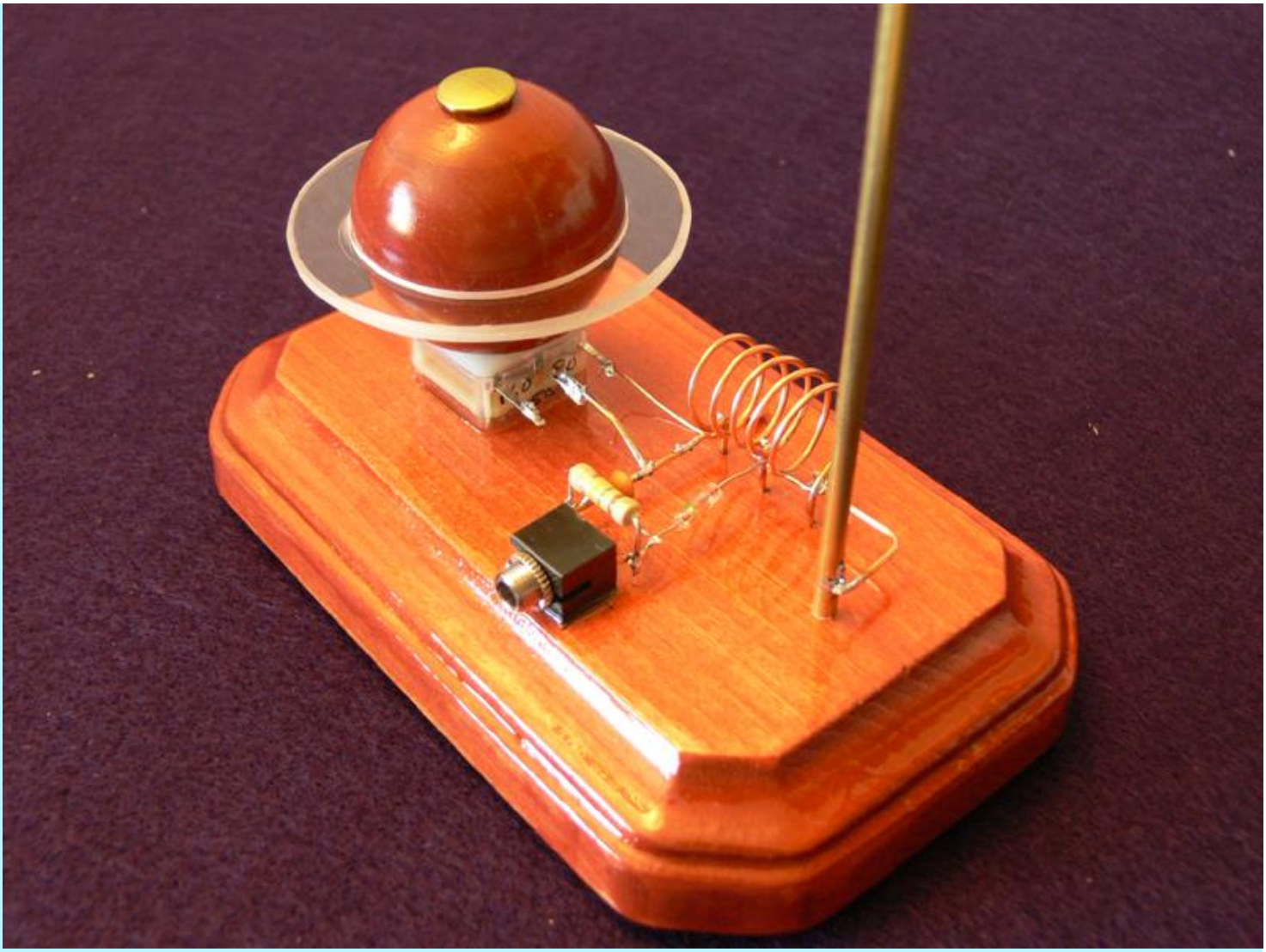
The vernier dial is large to accomodate ease of tuning, and the vernier makes it easy to separate stations. Two golden (brass) wood screws fix the face plate to the base. Holes for the face plate were made with special plastic drills, but ordinary drills may be used if drilled very SLOWLY. The knob is removable.

FM Crystal Circuit #2

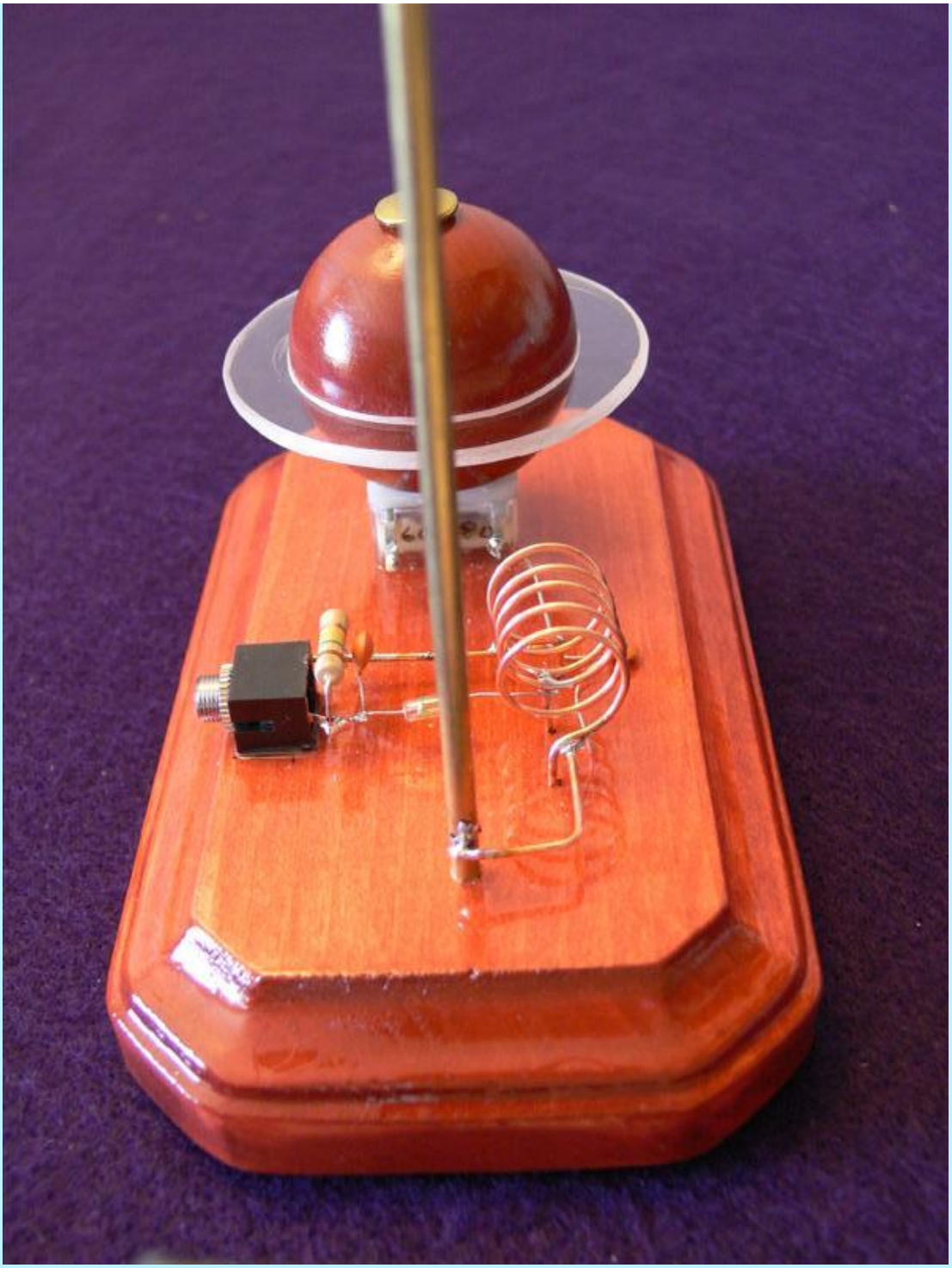


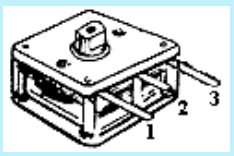
L - 5 turns AWG#18 bare copper or silver wire, 12mm inside diameter, tapped at 2.5 turns
D - 1N34 or rock crystal diode
C1 - 82 pf capacitor
C2 - 80 pf air variable capacitor
C3 - 18 pf capacitor
R - 150K resistor

The following photographs show the circuit wired with the handmade Saturn Dial. and knob. It is perhaps not as visually striking as set No. 1, but it works just as well. In fact, this set was the original version. Notice that all the wiring and coil are copper.



The Saturn dial and knob were fashioned from a "doll's head" from Michael's Arts and Crafts, a piece of lucite cut with two circle cutters, and a brass paper fastener. The knob is fixed to the tuning capacitor with a small machine screw that fits in the hole below the brass fastener. The most difficult part of this was fashioning "Saturn's rings". This must be done very carefully and slowly. The inside edge should be cut slightly undersized and then sanded with a drum sander to fit snugly. The outside edges can be sanded with fine sandpaper and polished with a plastic polisher.



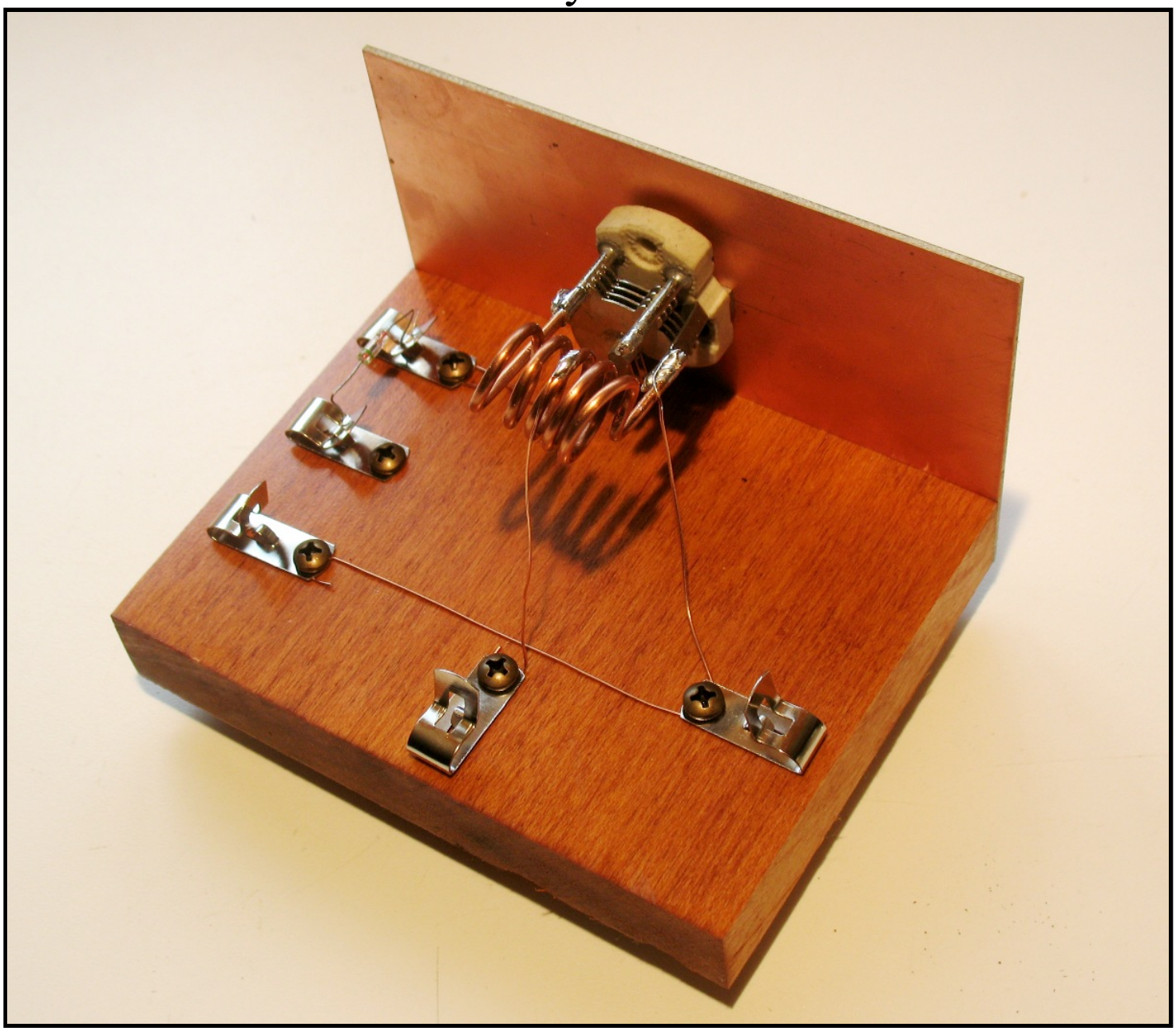


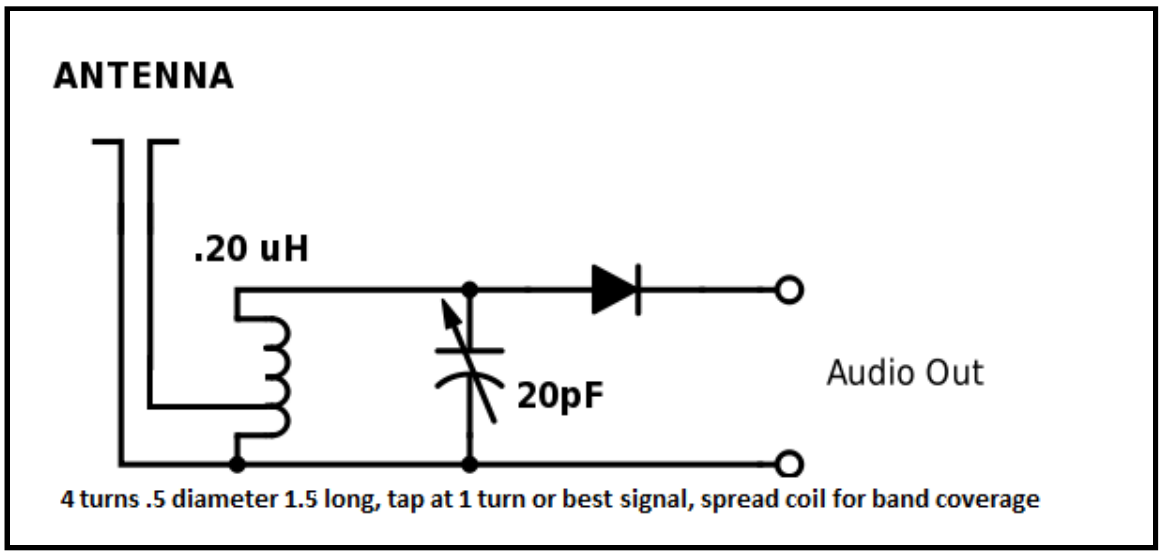
The air variable capacitor may be obtained from Electronix Express at http://www.elexp.com/. Part number 14VCRF10-280P. The 80 pf side is recommended for the second circuit, contacts 2-3. Contacts 1 and 3 were used for the first circuit (50pf).

- OSC: 5-59 pf
- ANT: 5-142 pf
- OSC and ANT Trimmer 10pf range

Home -- If you want to contact me go to the home page and go through the Music Clinic link.

FM Crystal Set





By slope detection you can get FM. If you live close to the stations you should do well with Hi Z phones. I use an audio amp, I am far from the stations. I get three.

Another portable FM crystal radio

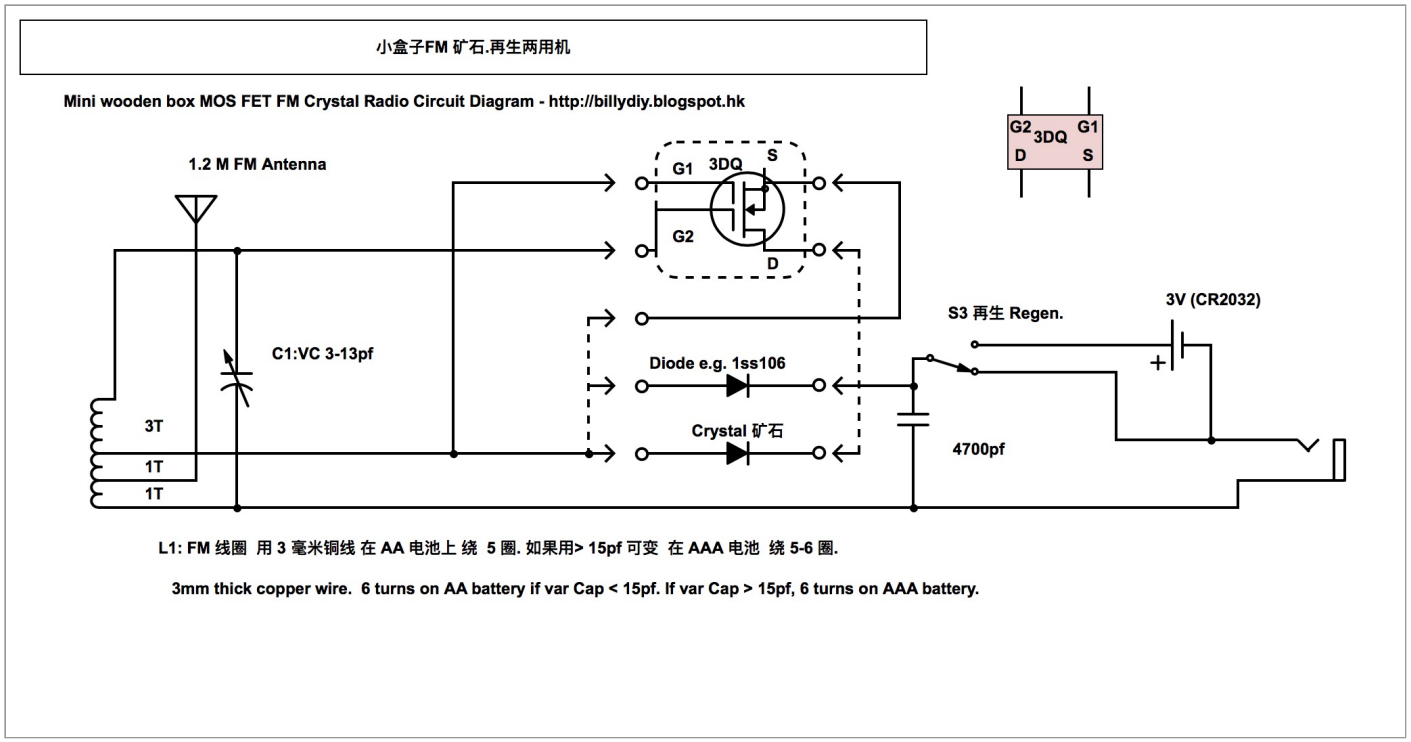
cheungbx

Another portable FM crystal radio

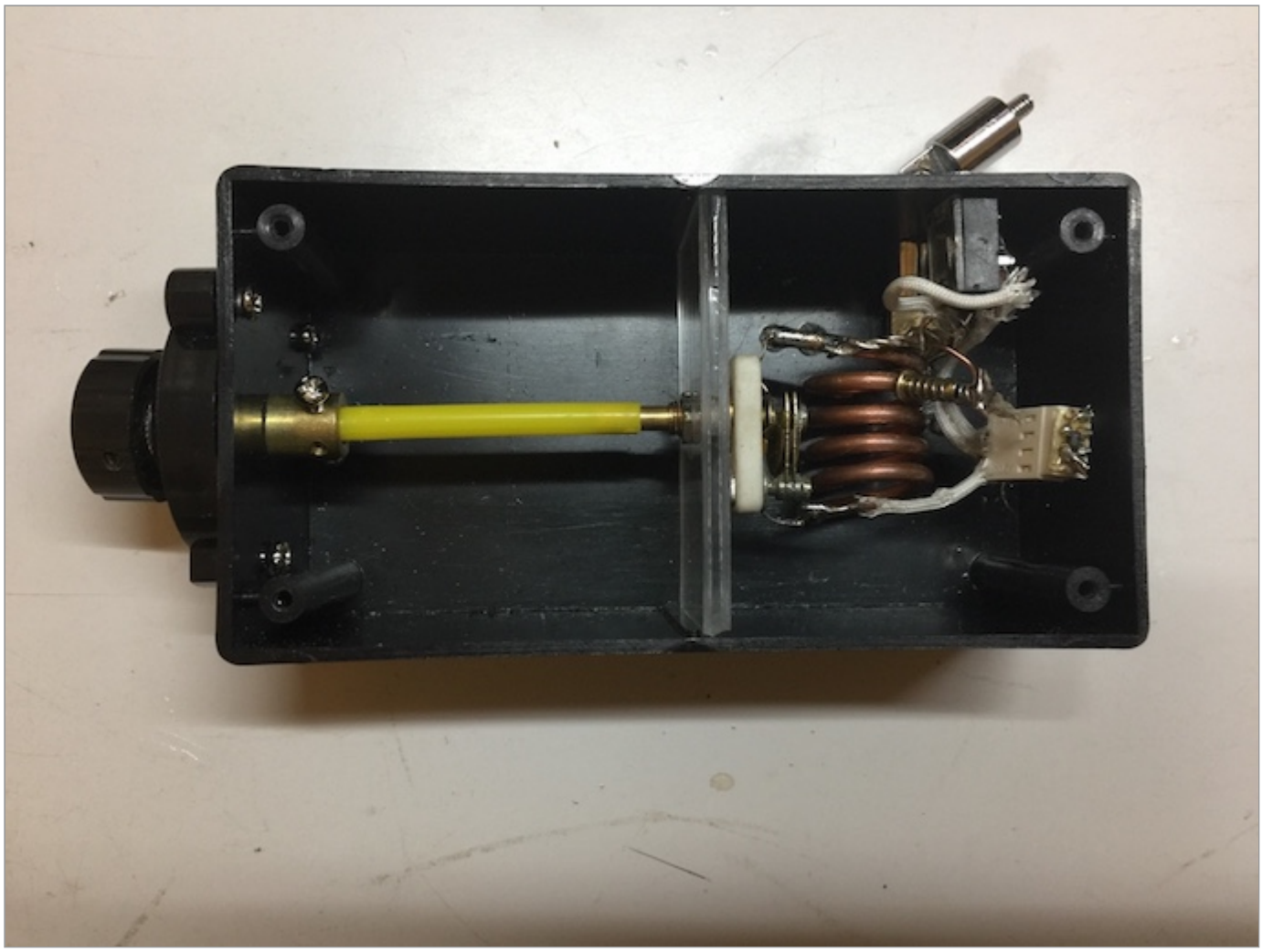
Fri Sep 08, 2017 4:16 pm

I am going to visit southern Taiwan tomorrow.
Made another portable FM crystal radio to test there.
Tested ok in HK, with clear and loud enough output for most FM stations in HK.
Made a clamp to hold the tab in place, while adjusting.

ATTACHMENTS



Circuit Diagram.jpg (202.37 KiB) Viewed 282 times



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IMG_4152.JPG (106.58 KiB) Viewed 281 times

IMG_4153.JPG (97.09 KiB) Viewed 281 times

Cheers!
Billy

billydiy.blogspot.hk
My dream is to make crystal radios as easy to use as regular radios.

andreaconsole

Re: Another portable FM crystal radio

Sun Sep 10, 2017 1:19 pm

Every time a new masterpiece!
Did you obtain any improvements with this new schematics?

cheungbx

Re: Another portable FM crystal radio

Tue Sep 12, 2017 1:41 pm

Just tested the FM Crystal Radio (without Battery) in Taiwan, Kao Hsiung.
It's loud and clear and I recorded a testing radio.

Antenna Technology

Zana Lane

First Edition, 2012

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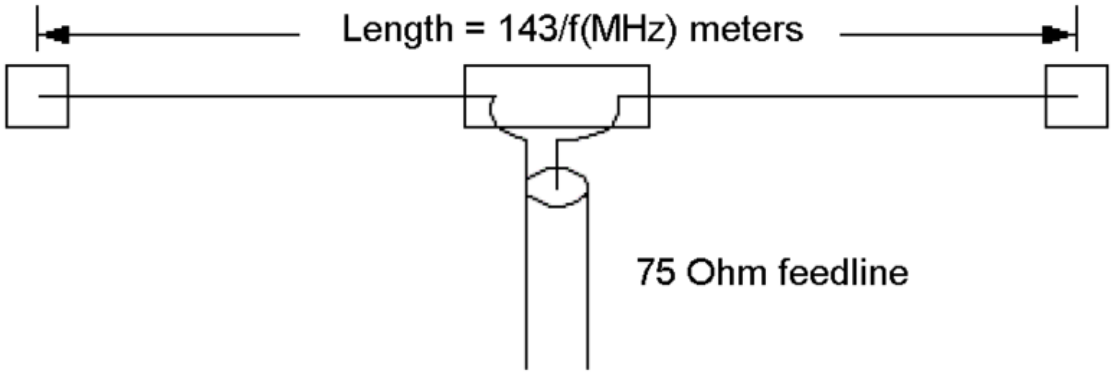
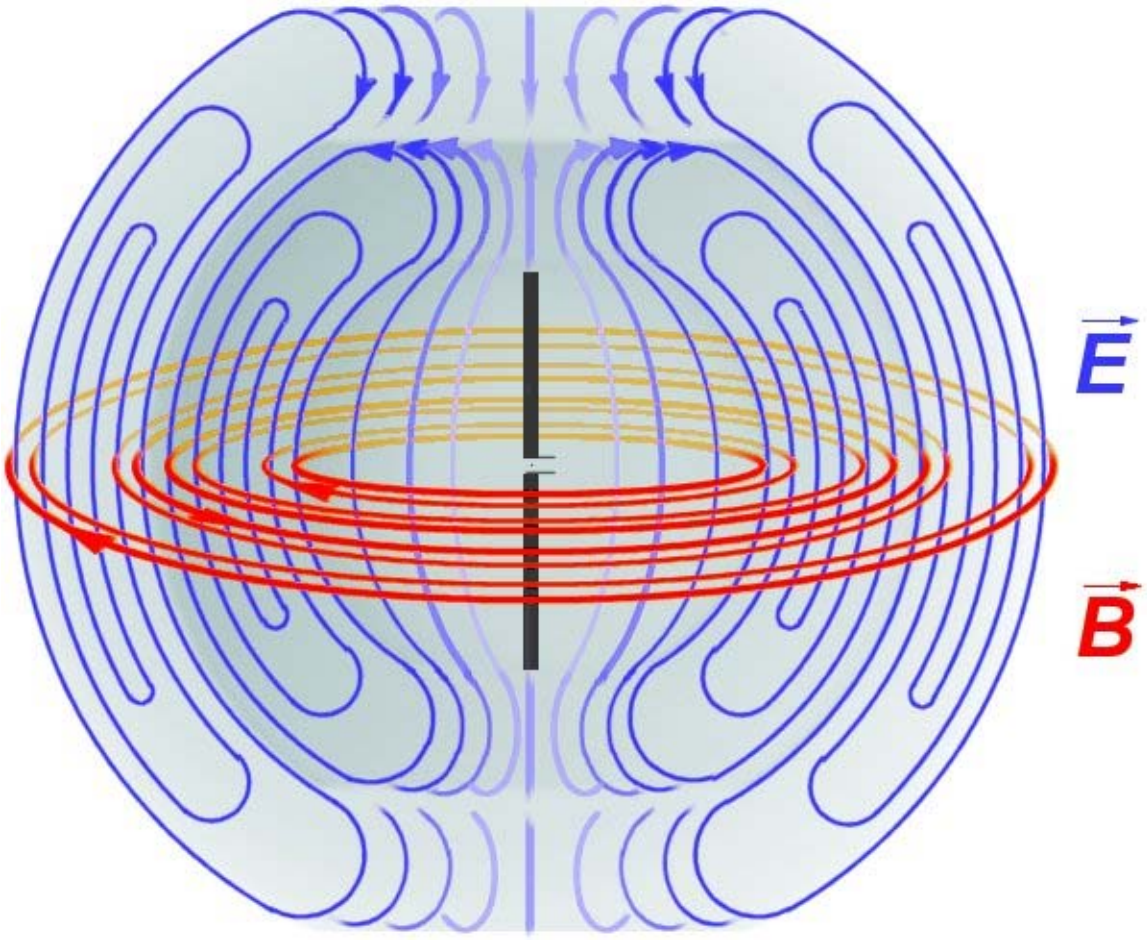
Chapter 1

Dipole Antenna



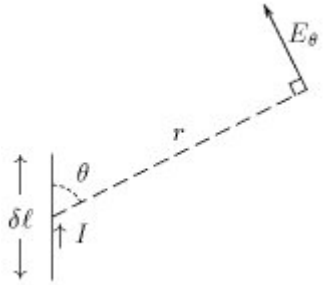
A schematic of a half-wave dipole antenna that a shortwave listener might build.

A **dipole antenna** is a radio antenna that can be made of a simple wire, with a center-fed driven element. It consists of two metal conductors of rod or wire, oriented parallel and collinear with each other (in line with each other), with a small space between them. The radio frequency voltage is applied to the antenna at the center, between the two conductors. These antennas are the simplest practical antennas from a theoretical point of view. They are used alone as antennas, notably in traditional "rabbit ears" television antennas, and as the driven element in many other types of antennas, such as the Yagi. Dipole antennas were invented by German physicist Heinrich Hertz around 1886 in his pioneering experiments with radio waves.



Electric fields *(blue)* and magnetic fields *(red)* radiated by a dipole antenna

Elementary doublet



An elementary doublet is a small length of conductor $\delta\ell$(small compared to the wavelength λ) carrying an alternating current:

$$I = I_m e^{i\omega t}.$$

Here $\omega = 2\pi f$ is the angular frequency (and f the frequency), and i is $\sqrt{-1}$, so that I is a phasor.

Note that this dipole cannot be physically constructed because the current needs somewhere to come from and somewhere to go to. In reality, this small length of conductor will be just one of the multiple segments into which we must divide a real antenna, in order to calculate its properties. The interest of this imaginary elementary antenna is that we can easily calculate the electrical far field of the electromagnetic wave radiated by each elementary doublet. We give just the result:

$$E_\theta = \frac{-i I_o \sin\theta}{2\varepsilon_o c r} \frac{\delta\ell}{\lambda} e^{i(\omega t - kr)}$$

Where,

- E_θ is the far electric field of the electromagnetic wave radiated in the θ direction.
- ε_o is the permittivity of vacuum.
- c is the speed of light in vacuum.
- r is the distance from the doublet to the point where the electrical field E_θ is evaluated.
- k is the wavenumber $k = \frac{2\pi}{\lambda}$

The exponent of e accounts for the phase dependence of the electrical field on time and the distance from the dipole.

The far electric field E_θ of the electromagnetic wave is coplanar with the conductor and perpendicular with the line joining the dipole to the point where the field is evaluated. If the dipole is placed in the center of a sphere in the axis south-north, the electric field would be parallel to geographic meridians and the magnetic field of the electromagnetic wave would be parallel to geographic parallels.

Near Field

The above formulas are valid for the far field of the antenna ($r \gg \lambda/(2\pi)$), and are the only contribution to the radiated field. The formulas in the near field have additional terms that reduce with r^2 and r^3. These are,

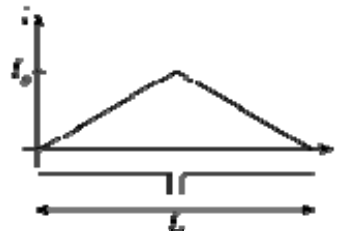
$$E_r = \frac{Z}{2\pi} I_0 \delta l \left(\frac{1}{r^2} - i \frac{\lambda}{2\pi r^3} \right) e^{i(\omega t - kr)} \cos(\theta)$$

$$E_\theta = i \frac{Z}{2\lambda} I_0 \delta l \left(\frac{1}{r} - i \frac{\lambda}{2\pi r^2} - \frac{\lambda^2}{4\pi^2 r^3} \right) e^{i(\omega t - kr)} \sin(\theta)$$

$$H_\phi = i \frac{1}{2\lambda} I_0 \delta l \left(\frac{1}{r} - i \frac{\lambda}{2\pi r^2} \right) e^{i(\omega t - kr)} \sin(\theta)$$

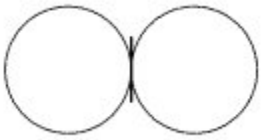
where $Z = \sqrt{\mu/\varepsilon} = 1/(\varepsilon c) = \mu c$. The energy associated with the term of the near field flows back and forward out and into the antenna.

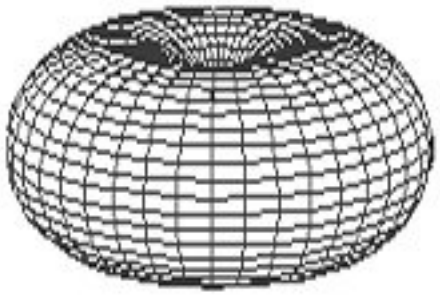
Short dipole



A short dipole is a physically feasible dipole formed by two conductors with a total length Lvery small compared with the wavelength λ. The two conducting wires are fed at the centre of the dipole. We assume the hypothesis that the current is maximal at the centre (where the dipole is fed) and that it decreases linearly to be zero at the ends of the wires. Note that the direction of the current is the same in both the dipole branches - to the right in both or to the left in both. The far field $E_\theta$of the electromagnetic wave radiated by this dipole is:

$$E_\theta = \frac{-iI_0 \sin\theta}{4\varepsilon_0 cr}\frac{L}{\lambda}e^{i(\omega t - kr)}.$$





Emission is maximal in the plane perpendicular to the dipole and zero in the direction of wires which is the direction of the current. The emission diagram is circular section torus shaped (right image) with zero inner diameter. In the left image the doublet is vertical in the torus centre.

Knowing this electric field, we can compute the total emitted power and then compute the resistive part of the series impedance of this dipole due to the radiated field, known as the radiation resistance:

$$R_{series} = \frac{\pi}{6} Z_0 \left(\frac{L}{\lambda}\right)^2 \text{ (for } L \ll \lambda\text{)}.$$

where Z_0 is the impedance of free space. Using a common approximation of $Z_0 = 120\pi$ ohms, we get:

$$R_{series} = 20\pi^2 \left(\frac{L}{\lambda}\right)^2 \text{ ohms}$$

Antenna gain

Antenna gain is the ratio of surface power radiated by the antenna to the surface power radiated by a hypothetical isotropic antenna:

$$G = \frac{\left(\frac{P}{S}\right)_{ant}}{\left(\frac{P}{S}\right)_{iso}}$$

The surface power carried by an electromagnetic wave is:

$$\left(\frac{P}{S}\right)_{ant} = \frac{1}{2}\varepsilon_0 c E_\theta^2 \simeq \frac{1}{120\pi} E_\theta^2$$

The surface power radiated by an isotropic antenna feed with the same power is:

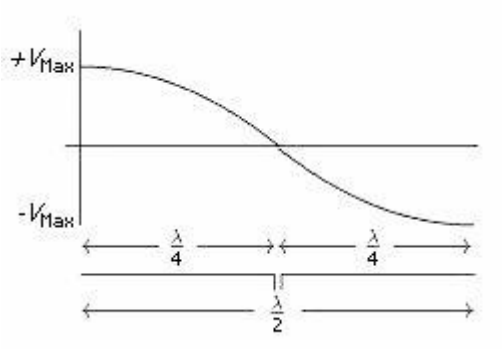
$$\left(\frac{P}{S}\right)_{iso} = \frac{\frac{1}{2} R_{series} I_0^2}{4\pi r^2}$$

Substituting values for the case of a short dipole, final result is:

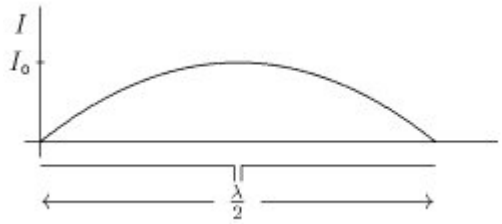
$$G = \frac{\pi \left(\frac{L}{\lambda}\right)^2}{\varepsilon_0 c \frac{2\pi}{3\varepsilon_0 c} \left(\frac{L}{\lambda}\right)^2} = 1.5 = 1.76 \text{ dBi}$$

dBi simply means decibels gain, relative to an **i**sotropic antenna.

Half-wave antenna



Typically a dipole antenna is formed by two quarter wavelength conductors or elements placed back to back for a total length of $\lambda/2$. A standing wave on an element of a length $\sim\lambda/4$ yields the greatest voltage differential, as one end of the element is at a node while the other is at an antinode of the wave. The larger the differential voltage, the greater the current between the elements.



Assuming a sinusoidal distribution, the current impressed by this voltage differential is given by:

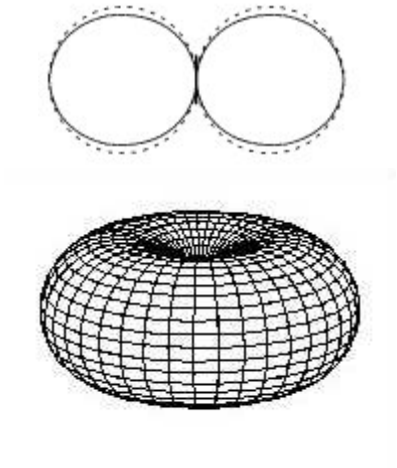
$$I = I_{\circ} e^{i\omega t} \cos k\ell$$

For the far-field case, the formula for the electric field of a radiating electromagnetic wave is somewhat more complex:

$$E_{\theta} = \frac{-iI_{\circ}}{2\pi\varepsilon_{\circ} cr} \frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta} e^{i(\omega t - kr)}$$

But the fraction $\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}$ is not very different from $\sin\theta$.

The resulting emission diagram is a slightly flattened torus.



The image on the left shows the section of the emission pattern. We have drawn, in dotted lines, the emission pattern of a short dipole. We can see that the two patterns are very similar. The image at right shows the perspective view of the same emission pattern.

This time it is not possible to compute analytically the total power emitted by the antenna (the last formula does not allow), though a simple numerical integration or series expansion leads to the more precise, actual value of the half-wave resistance:

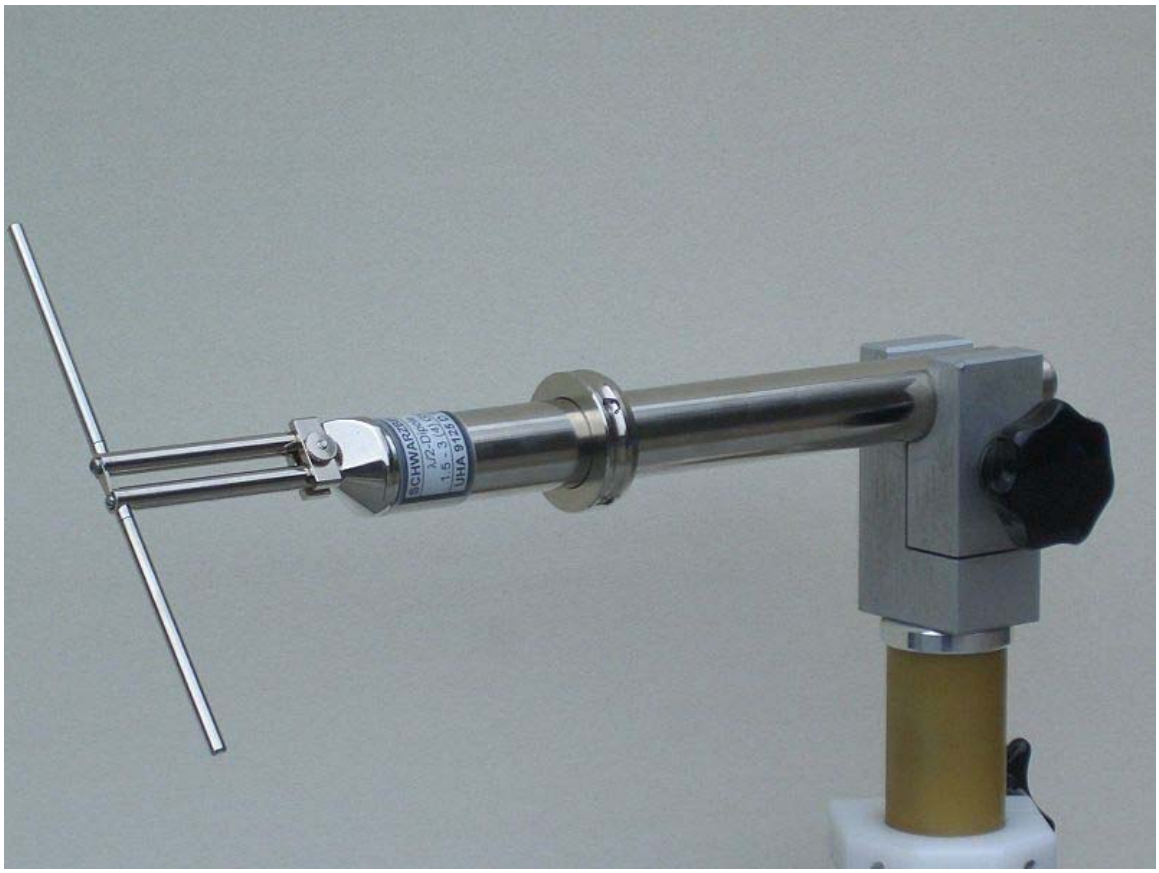
$$R_{\frac{\lambda}{2}} = 60\mathrm{Cin}(2\pi) = 60\left[\ln(2\pi\gamma) - \mathrm{Ci}(2\pi)\right] = 120\int_0^{\frac{\pi}{2}} \frac{\cos\left(\frac{\pi}{2}\cos\theta\right)^2}{\sin\theta}d\theta,$$
$$= 15\left[2\pi^2 - \frac{1}{3}\pi^4 + \frac{4}{135}\pi^6 - \frac{1}{630}\pi^8 + \frac{4}{70875}\pi^{10} \ldots - (-1)^n\frac{(2\pi)^{2n}}{n(2n)!}\right],$$
$$\approx 73.13\Omega;$$

This leads to the gain of a dipole antenna, $G_{\frac{\lambda}{2}}$:

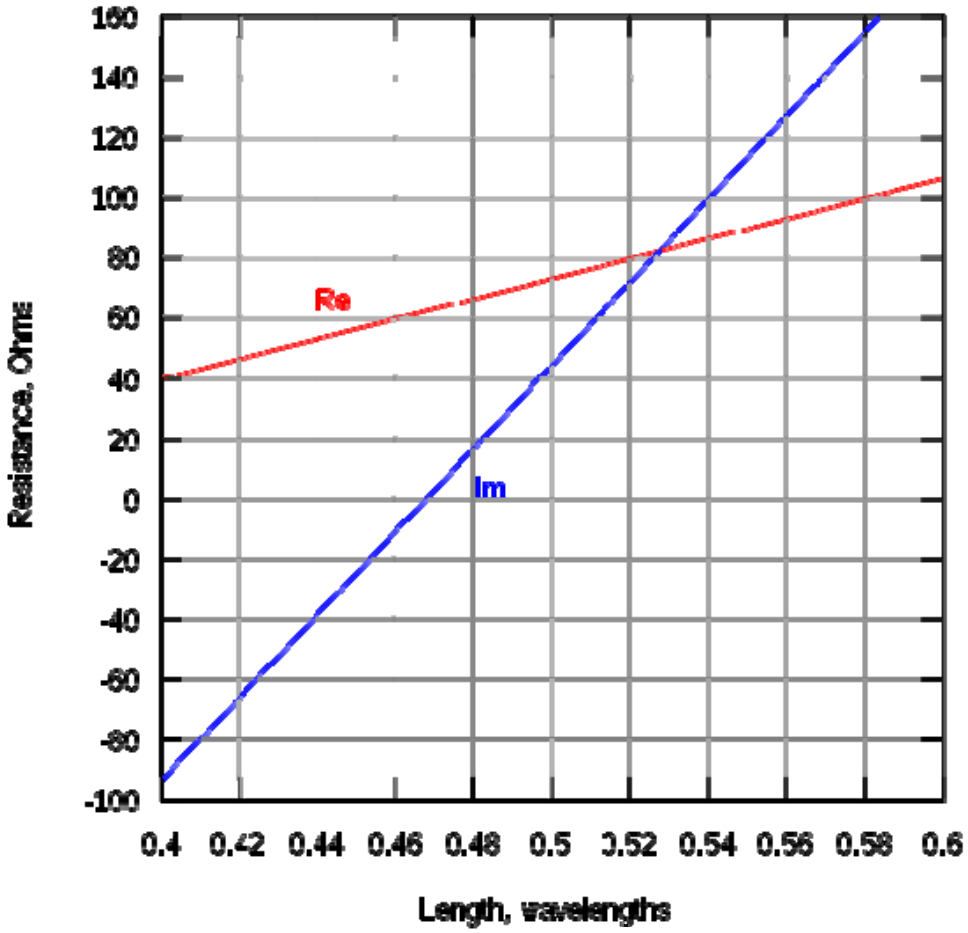
$$G_{\frac{\lambda}{2}} = \frac{60^2}{30R_{\frac{\lambda}{2}}} = \frac{3600}{30R_{\frac{\lambda}{2}}} = \frac{120}{R_{\frac{\lambda}{2}}} = \frac{1}{\int_0^{\frac{\pi}{2}} \frac{\cos\left(\frac{\pi}{2}\cos\theta\right)^2}{\sin\theta}d\theta},$$
$$\approx \frac{120}{73.1296} \approx 1.64 \approx 2.15\,\mathrm{dBi};$$

The resistance, however, is not enough to characterize the dipole impedance, as there is also an imaginary part——it is better to measure the impedance.

In the image below, the real and imaginary parts of a dipole's impedance are drawn for lengths going from $0.4\ \lambda$ to $0.6\ \lambda$, accompanied by a chart comparing the gains of dipole antennas of other lengths (note that gains are **not** in dBi but in natural number):



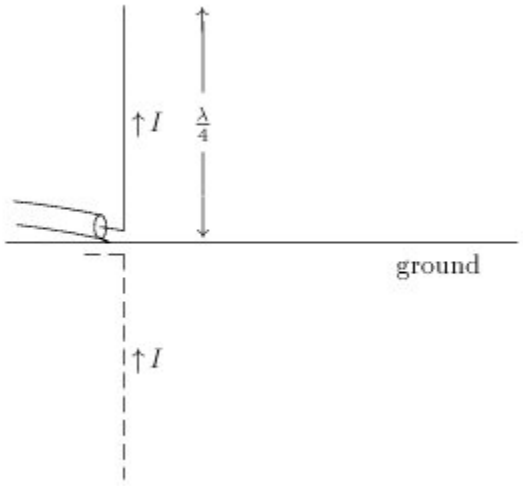
UHF–Half–Wave Dipole, 1.0–4 GHz



Gain of dipole antennas

length **L** in λ	Gain	Gain(dB)
≪0.1	1.50	1.76dB
0.5	**1.64**	**2.15dB**
1.0	1.80	2.55dB
1.5	2.00	3.01dB
2.0	2.30	3.62dB
3.0	2.80	4.47dB
4.0	3.50	5.44dB
8.0	7.10	8.51dB

Quarter-wave antenna



The antenna and its image form a $\frac{\lambda}{2}$dipole that radiates only upward.

The quarter wave monopole antenna is a single element antenna fed at one end, that behaves as a dipole antenna. It is formed by a conductor $\frac{\lambda}{4}$ in length. It is fed in the lower end, which is near a conductive surface which works as a reflector. The current in the reflected image has the same direction and phase as the current in the real antenna. The quarter-wave conductor and its image together form a half-wave dipole that radiates only in the upper half of space.

In this upper side of space the emitted field has the same amplitude of the field radiated by a half-wave dipole fed with the same current. Therefore, the total emitted power is one-half the emitted power of a half-wave dipole fed with the same current. As the current is the same, the radiation resistance (real part of series impedance) will be one-half of the series impedance of a half-wave dipole. As the reactive part is also divided by 2, the impedance of a quarter wave antenna is $\frac{73+i43}{2}=36+i21$ohms. Since the fields above ground are the same as for the dipole, but only half the power is applied, the gain is twice (3dB over) that for a half-wave dipole ($\frac{\lambda}{2}$), that is 5.14 dBi.

The earth can be used as ground plane, but it is a poor conductor: the reflected antenna image is only clear at glancing angles (far from the antenna). At these glancing angles, electromagnetic fields and radiation patterns are thus the same as for a half-wave dipole.

Naturally, the impedance of the earth is far inferior to that of a good conductor ground plane -- this can be improved (at cost) by laying a copper mesh.

When ground is not available (such as in a vehicle) other metallic surfaces can serve as a ground plane (typically the vehicle's roof). Alternatively, radial wires placed at the base of the antenna can simulate a ground plane. For VHF bands, the radiating and ground-plane elements can be constructed from rigid rods or tubes.

Dipole characteristics

Frequency versus length

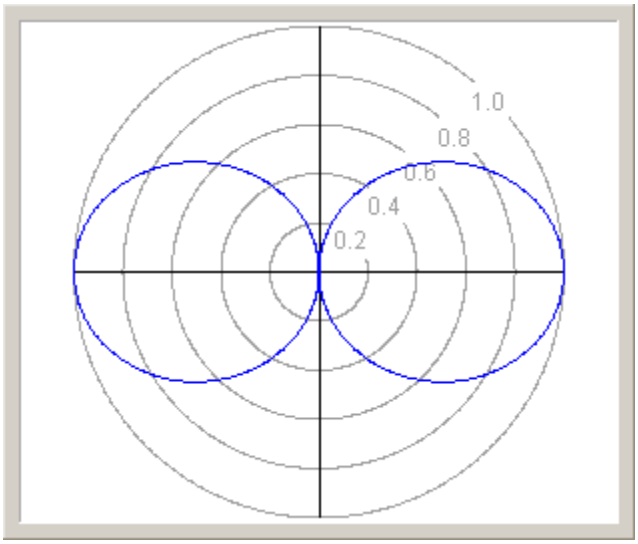
Dipoles that are much smaller than the wavelength of the signal are called *Hertzian, short, or infinitesimal dipoles*. These have a very low radiation resistance and a high reactance, making them inefficient, but they are often the only available antennas at very long wavelengths. Dipoles whose length is half the wavelength of the signal are called *half-wave dipoles*, and are more efficient. In general radio engineering, the term *dipole* usually means a half-wave dipole (center-fed).

A half-wave dipole is cut to length l for frequency f MHz according to the formula $l = \frac{143}{f}$ where l is in metres or $l = \frac{468}{f}$ where l is in feet. This is because the impedance of the dipole is resistive pure at about this length. The length of the dipole antenna is about 95% of half a wavelength at the speed of light in free space.

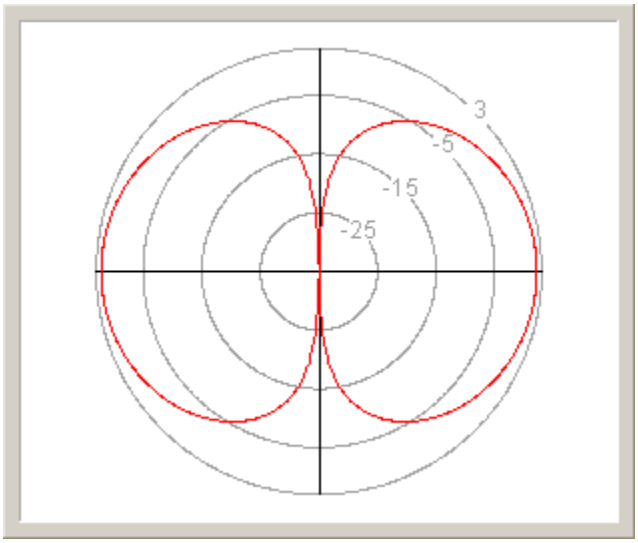
The magic numbers above are derived from a one Hz wavelength which is the distance that light radio travels in one second. Speed of light in vacuum is 299,792,458 m/s, which is divided by 1 million to account for MHz rather than Hz, which is then divided by 2 for a half-wave dipole antenna. A fudge factor of approximately 0.95 is multiplied to account for the damping due to radiation, which results in the magic number of 143 m·MHz or 468 ft·MHz.

Radiation pattern and gain

Dipoles have an omnidirectional radiation pattern, shaped like a toroid (doughnut) symmetrical about the axis of the dipole. The radiation is maximum at right angles to the dipole, dropping off to zero on the antenna's axis. The theoretical maximum gain of a Hertzian dipole is 10 log 1.5 or 1.76 dBi. The maximum theoretical gain of a λ/2-dipole is 10 log 1.64 or 2.15 dBi.



Radiation pattern of a half-wave dipole antenna. The scale is linear.



Gain of a half-wave dipole (same as left). The scale is in dBi (decibels over isotropic).

Feeder line

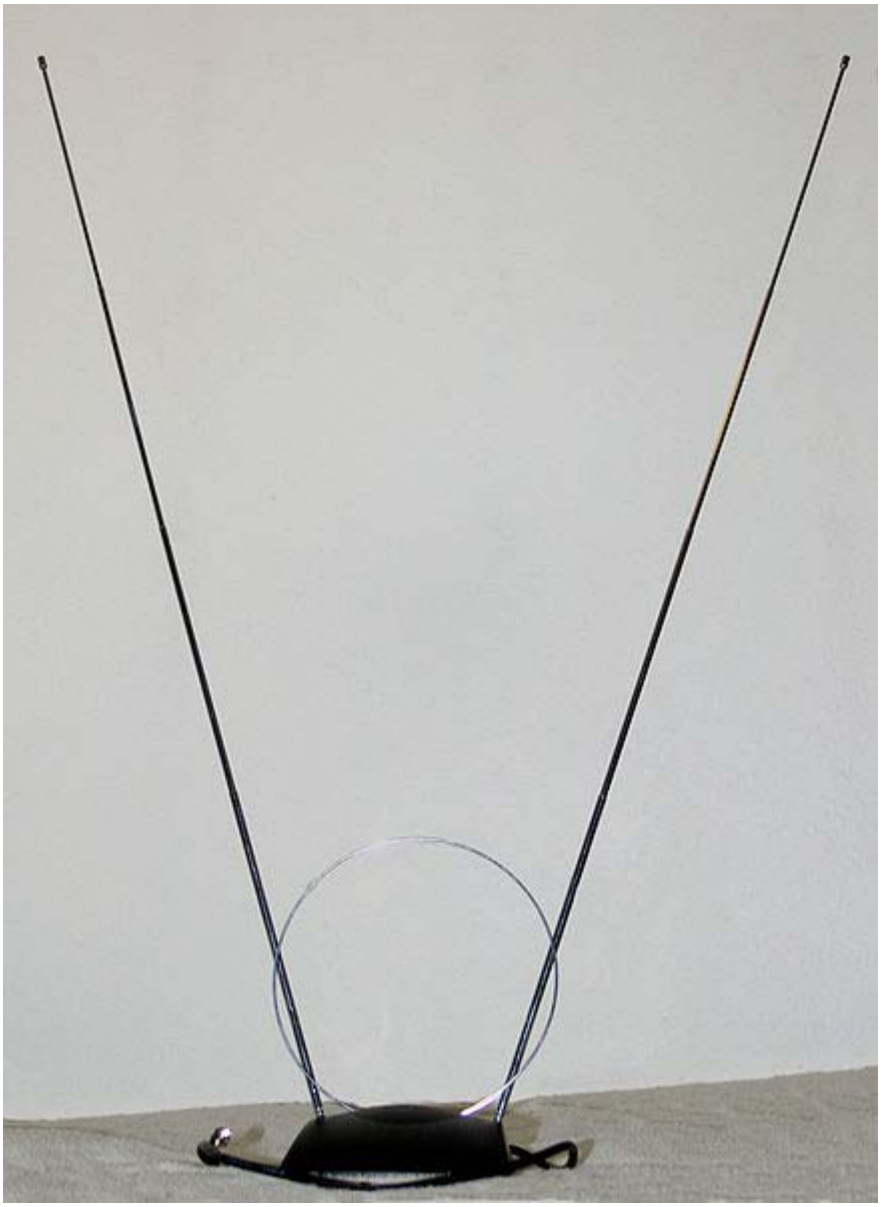
Ideally, a half-wave (λ/2) dipole should be fed with a balanced line matching the theoretical 73 ohm impedance of the antenna. A folded dipole uses a 300 ohm balanced feeder line.

Many people have had success in feeding a dipole directly with a coaxial cable feed rather than a ladder-line. However, coax is not symmetrical and thus not a balanced feeder. It is unbalanced, because the outer shield is connected to earth potential at the other end. When a balanced antenna such as a dipole is fed with an unbalanced feeder, common mode currents can cause the coax line to radiate in addition to the antenna itself,

and the radiation pattern may be asymmetrically distorted. This can be remedied with the use of a balun.

Common applications

Set-top TV antenna



A "rabbit-ears" antenna with a UHF loop antenna.

The most common dipole antenna is the type used with televisions, often colloquially referred to as "**rabbit ears**" or "**bunny ears**." While in most applications the dipole elements are arranged along the same line, rabbit ears are adjustable in length and angle. Larger dipoles are sometimes hung in a V shape with the center near the radio equipment on the ground or the ends on the ground with the center supported. Shorter dipoles can be

hung vertically. Some have extra elements to get better reception such as loops (especially for UHF transmissions), which can be turnable around a vertical axis, or a dial, which modifies the electrical properties of the antenna at each dial position.

Folded dipole

Folded dipole antenna

Another common place one can see dipoles is as antennas for the FM band - these are folded dipoles. The tips of the antenna are folded back until they almost meet at the feedpoint, such that the antenna comprises one entire wavelength. This arrangement has a greater bandwidth than a standard half-wave dipole. If the conductor has a constant radius and cross-section, at resonance the input impedance is four times that of a half-wave dipole.

Shortwave antenna

A DIY-made dipole antenna with mast

Dipoles for longer wavelengths are made from solid or stranded wire. Portable dipole antennas are made from wire that can be rolled up when not in use. Ropes with weights on the ends can be thrown over supports such as tree branches and then used to hoist up the antenna. The center and the connecting cable can be hoisted up with the ends on the ground or the ends hoisted up between two supports in a V shape. While permanent antennas can be trimmed to the proper length, it is helpful if portable antennas are adjustable to allow for local conditions when moved. One easy way is to fold the ends of the elements to form loops and use adjustable clamps. The loops can then be used as attachment points.

It is important to fit a good insulator at the ends of the dipole, as failure to do so can lead to a flashover if the dipole is used with a transmitter. Various purchased or improvised insulators can be used.

Whip antenna

The whip antenna is probably the most common and simplest-looking antenna. These are monopoles, and the most common and practical is the quarter-wave monopole which could be considered as half of a dipole using a ground plane as the image of the other half. The commonly referred-to end-fed dipole is actually just a half-wave monopole whip antenna.

Dipoles versus whip antennas

Dipoles are generally more efficient than whip antennas (quarter-wave monopoles). The total radiated power and the radiation resistance are twice that of a quarter-wave monopole. Thus, if a whip antenna were used with an infinite perfectly conducting ground plane, then it would be as efficient in half-space as a dipole in free space an infinite distance from any conductive surfaces such as the earth's surface.

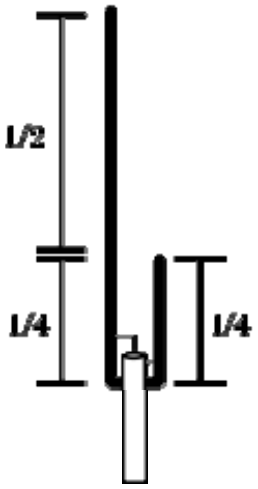
Dipole towers

Large constructed half-wavelength dipole towers include the Warsaw radio mast — the only half-wave dipole for longwave ever built — and Blaw-Knox Towers.

Military

US Military personnel occasionally use a 'doublet' antenna, especially during dismounted unconventional warfare. A radio operator may choose to bring several doublet antennas for different frequencies, such as an antenna cut to length for the set MEDEVAC (medical evacuation) frequency, NCS (net control station) frequency, and tactical frequency (the frequency used by troops in the field). This approach may not be acceptable depending on the mission. Note that a doublet antenna will not work with the standard SINCGARS radio when using frequency hopping(FH) but is effective for single channel (SC). A doublet antenna is more practical for radios not intended for FH.

Collinear antenna systems based on dipoles



J-Pole Antenna

Dipoles can be stacked end to end in phased arrays to make collinear antenna arrays, which exhibit more gain in certain directions—the toroidal radiation pattern is flattened out, giving maximum gain at right angles to the axis of the colinear array.

Slim Jim or J-pole

A Slim Jim or J-pole is a form of end-fed dipole connected to a quarter-wave stub matching section.

Dipole types

Ideal half-wavelength dipole

This type of antenna is a special case where each wire is exactly one-quarter of the wavelength, for a total of a half wavelength. The radiation resistance is about 73 ohms if wire diameter is ignored, making it easily matched to a coaxial transmission line. The directivity is a constant 1.64, or 2.15 dB. Actual gain will be a little less due to ohmic losses.

If the dipole is not driven at the centre then the feed point resistance will be higher. If the feed point is distance x from one end of a half wave ($\lambda/2$) dipole, the resistance will be described by the following equation.

$$R_r = \frac{75}{\sin^2\left(\frac{2\ \pi\ x}{\lambda}\right)}$$

If taken to the extreme then the feed point resistance of a $\lambda/2$ long rod is infinite, but it is possible to use a $\lambda/2$ pole as an aerial; the right way to drive it is to connect it to one

terminal of a parallel LC resonant circuit. The other side of the circuit must be connected to the braid of a coaxial cable lead and the core of the coaxial cable can be connected part way up the coil from the RF ground side. An alternative means of feeding this system is to use a second coil which is magnetically coupled to the coil attached to the aerial.

Folded dipole

A folded dipole is a half-wave dipole with an additional wire connecting its two ends. If the additional wire has the same diameter and cross-section as the dipole, two nearly identical radiating currents are generated. The resulting far-field emission pattern is nearly identical to the one for the single-wire dipole described above; however, at resonance its input (feedpoint) impedance R_{df} is four times the radiation resistance of a single-wire dipole. This is because for a fixed amount of power, the total radiating current I_0 is equal to twice the current in each wire and thus equal to twice the current at the feed point. Equating the average radiated power to the average power delivered at the feedpoint, we may write

$$\frac{1}{2}R_{\frac{\lambda}{2}}I_0^2 = \frac{1}{2}R_{fd}\left(I_0/2\right)^2 .$$

It follows that

$$R_{fd} = 4R_{\frac{\lambda}{2}} = 292.52\Omega.$$

The folded dipole is therefore well matched to 300-Ohm balanced transmission lines.

Hertzian dipole (current element)

The Hertzian dipole is a theoretical short dipole (significantly smaller than the wavelength) with a uniform current along its length. A true Hertzian dipole cannot physically exist, since the assumed current distribution implies an infinite accumulation of charge at its ends.

The radiation resistance is given by:

$$R_r = \frac{2\pi}{3}Z_0\left(\frac{\ell}{\lambda}\right)^2 .$$

where Z_0 is the impedance of free space. This is precisely four times the radiation resistance of the real short dipole with the linearly tapered current distribution.

The radiation resistance is typically a fraction of an ohm, making the infinitesimal dipole an inefficient radiator. The directivity D, which is the theoretical gain of the antenna assuming no ohmic losses (not real-world), is a constant of 1.5, which corresponds to

1.76 dB. Actual gain will be much less due to the ohmic losses and the loss inherent in connecting a transmission line to the antenna, which is very hard to do efficiently considering the incredibly low radiation resistance. The maximum effective aperture is:

$$A_e = \frac{3\lambda^2}{8\pi}$$

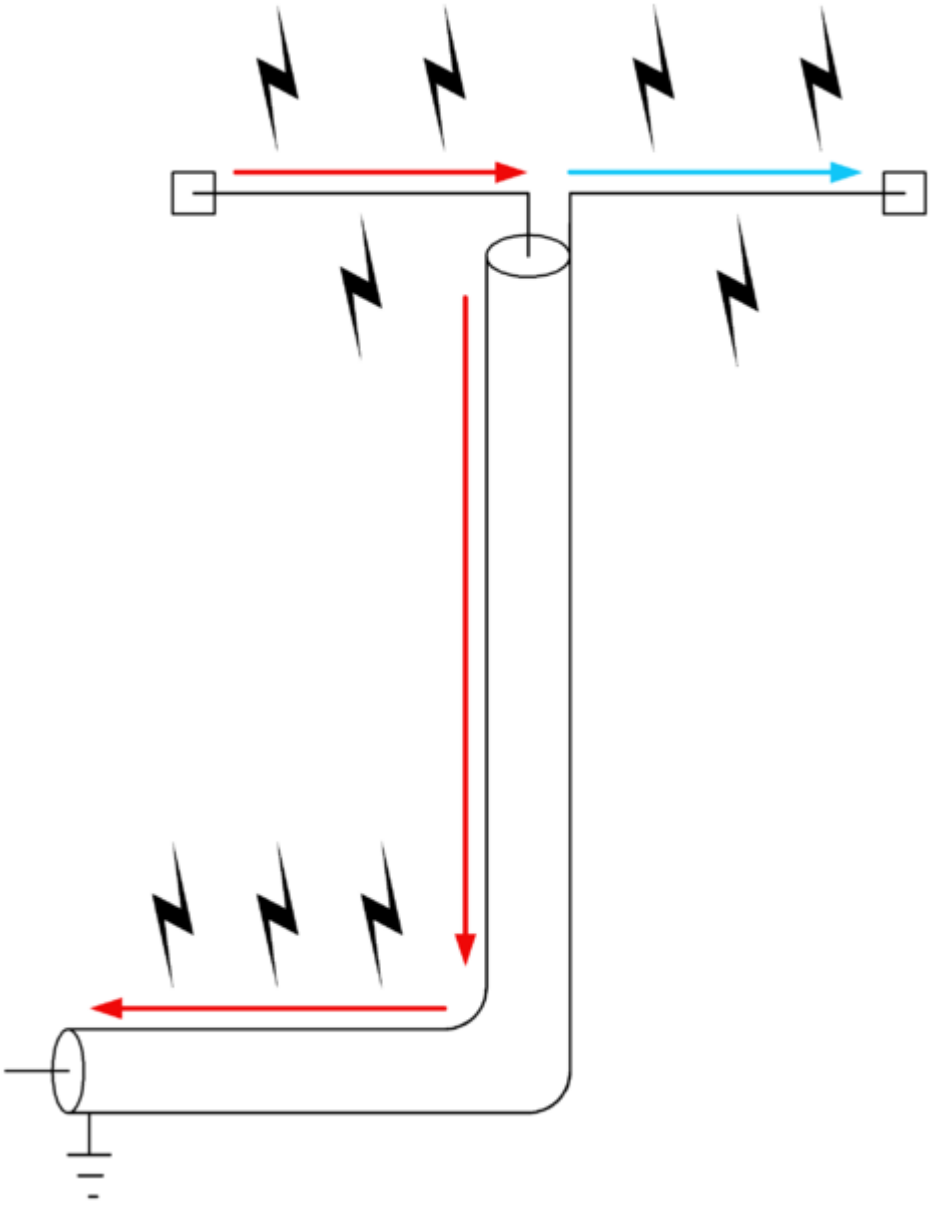
A surprising result is that even though the Hertzian dipole is minute, its effective aperture is comparable to antennas many times its size.

Dipole as a reference standard

Antenna gain is sometimes measured as "*x* dB above a dipole", which means that the antenna in question is being compared to a dipole, and has *x* dB more gain (has more directivity) than the dipole tuned to the same operating frequency. In this case one says the antenna has a gain of "*x* dBd". More often, gains are expressed relative to an isotropic radiator, which is an imaginary aerial that radiates equally in all directions. In this case one uses dBi instead of dBd. As it is impossible to build an isotropic radiator, gain measurements expressed relative to a dipole are more practical when a reference dipole aerial is used for experimental measurements. 0 dBd is often considered equal to 2.15 dBi.

From Babinet's principle, a dipole antenna is complementary to a slot antenna consisting of a slot the same size and shape as a dipole cut from an infinite sheet of metal; both give the same radiation pattern.

Dipole with baluns



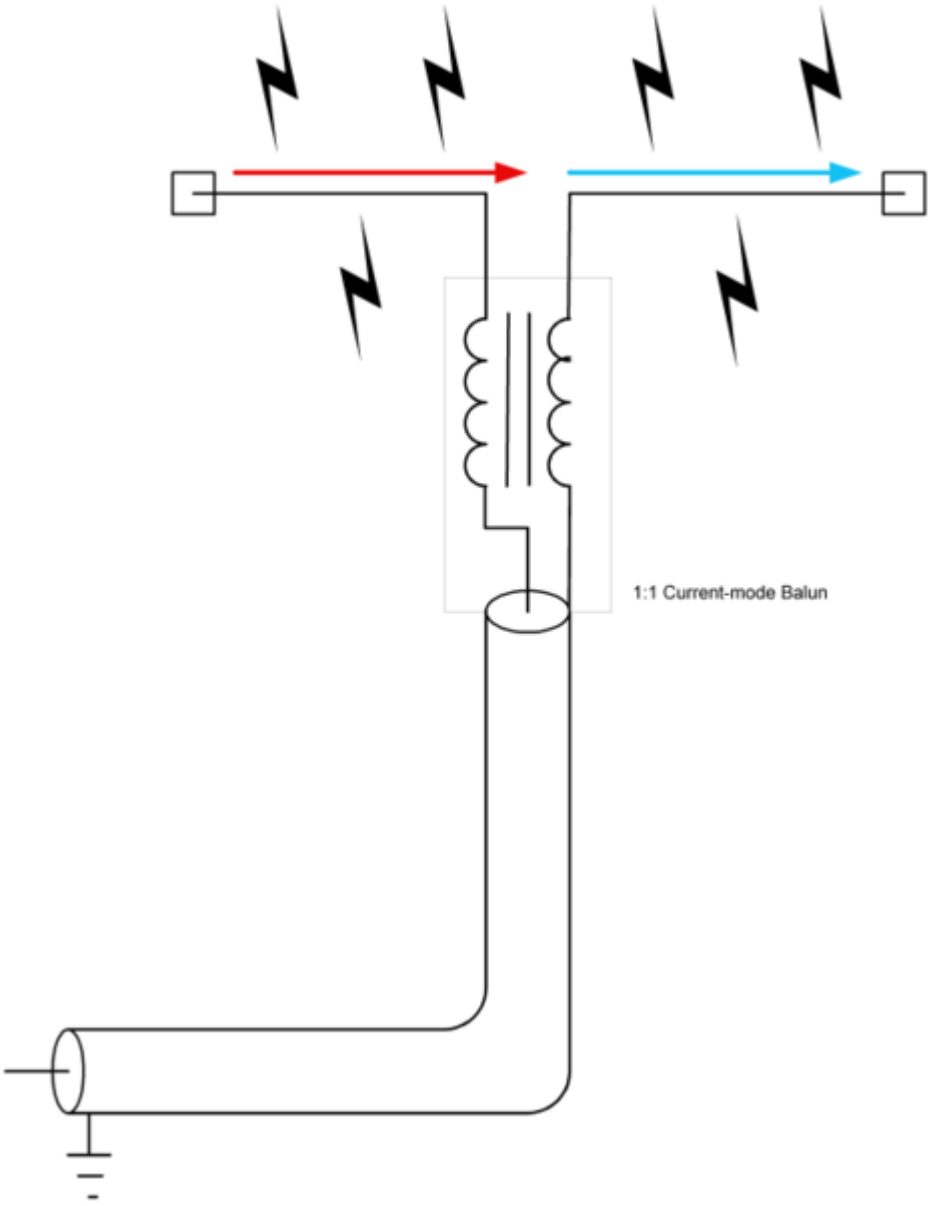
Coax and antenna both acting as radiators instead of only the antenna.

A dipole, being composed of two symmetrical ungrounded elements, works best when fed by a balanced transmission line, such as ladder line. When a dipole with an unbalanced feedline such as coaxial cable is used for transmitting, the shield side of the cable, in addition to the antenna, radiates. This can induce RF currents into other electronic equipment near the radiating feedline, causing RF interference. Furthermore, the antenna is not as efficient as it could be because it is radiating closer to the ground and its radiation (and reception) pattern may be distorted asymmetrically. At higher frequencies, where the length of the dipole becomes significantly shorter than the diameter of the feeder coax, this becomes a more significant problem. To prevent this, dipoles fed by coaxial cables have a balun between the cable and the antenna, to convert

the unbalanced signal provided by the coax to a balanced symmetrical signal for the antenna.

Several type of baluns are commonly used to transmit on a dipole: current baluns and coax baluns.

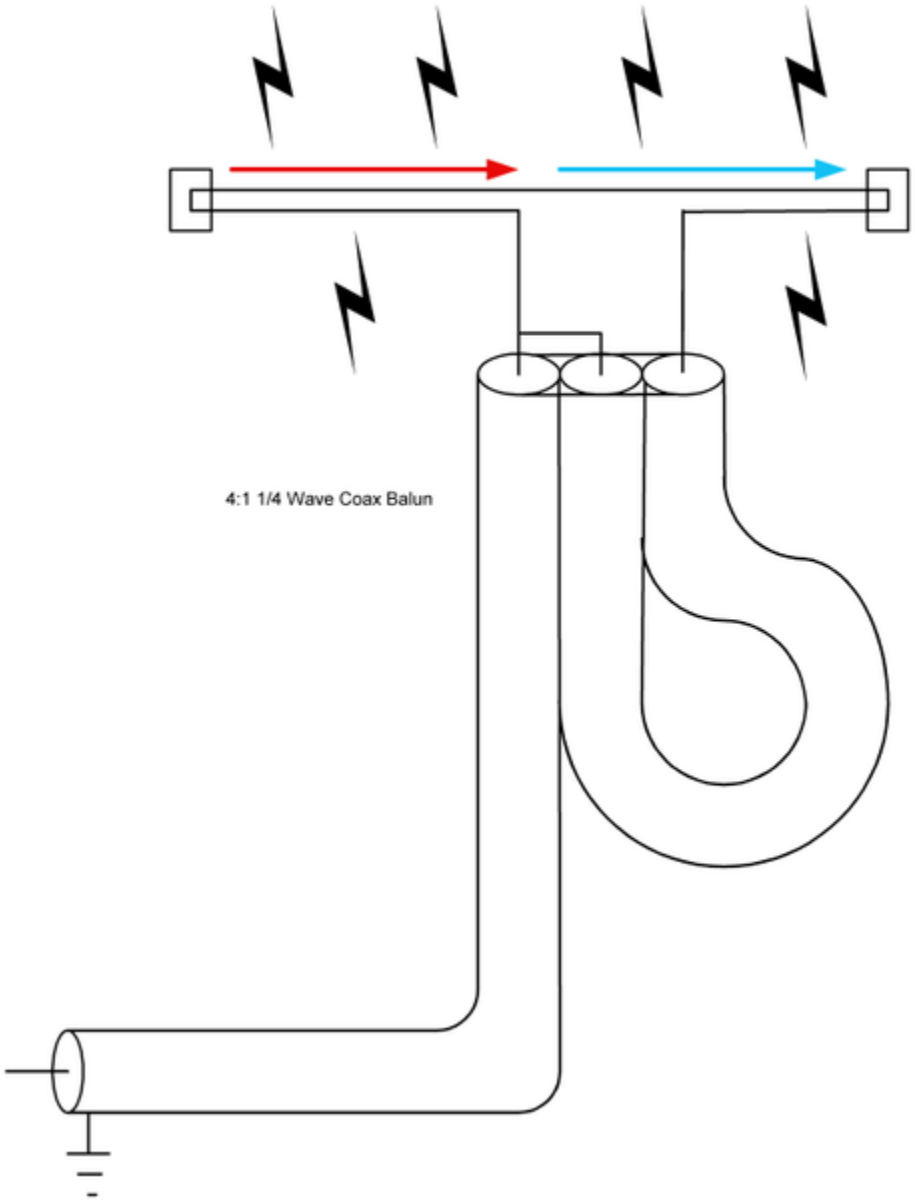
Current balun



Dipole with a current balun.

A current balun is a bit more expensive but has the characteristic of being more broadband. It can also be as simple as winding the coax cable over a ferrite core. Or nothing but coax cable:

Coax balun

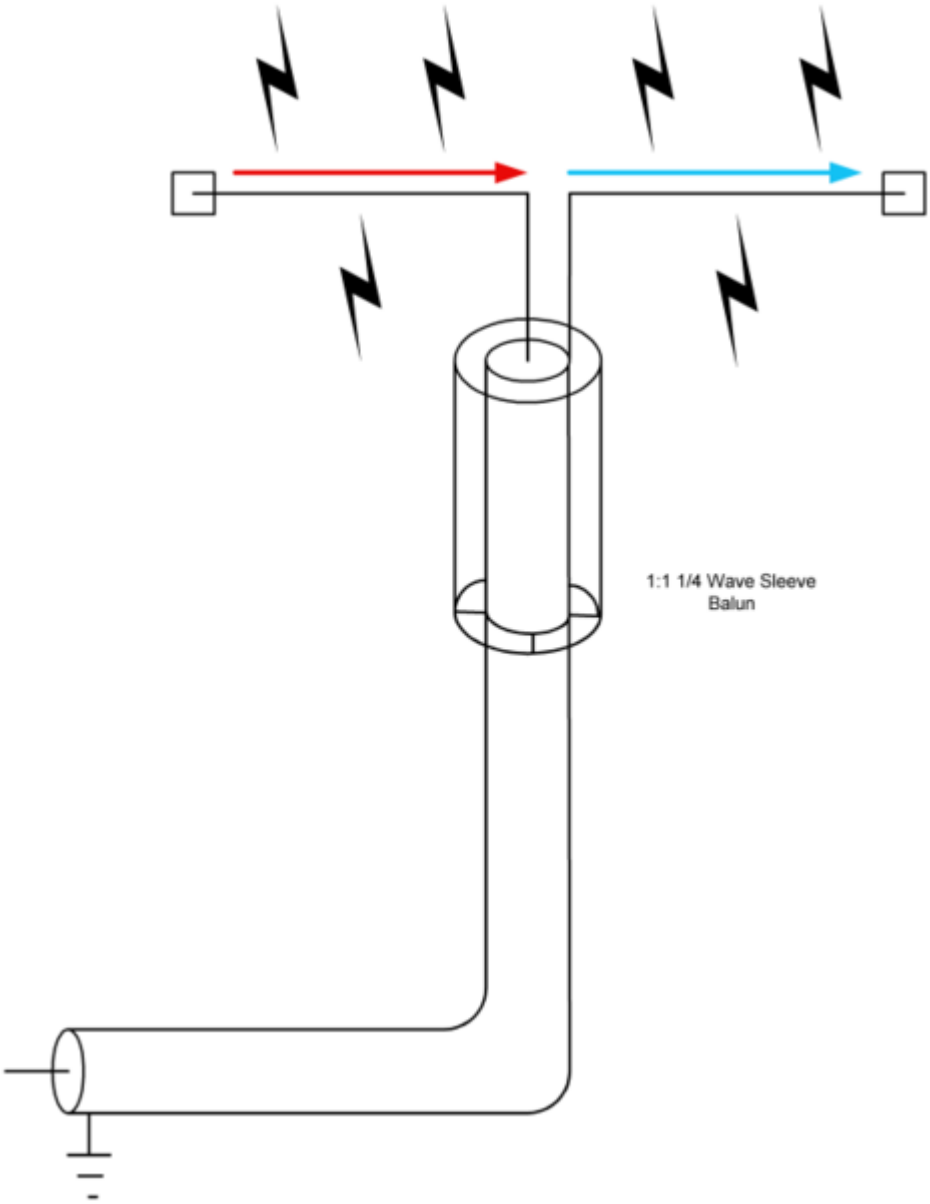


Here is a dipole using a coax balun.

A coax balun is a cost effective method to eliminate feeder radiation, but is limited to a narrow set of operating frequencies.

- One easy way to make a balun is a (λ/2) length of coaxial cable. The inner core of the cable is linked at each end to one of the balanced connections for a feeder or dipole. One of these terminals should be connected to the inner core of the coaxial feeder. All three braids should be connected together. This then forms a 4:1 balun which works correctly at only a narrow band of frequencies.

Sleeve balun



Here is a dipole using a sleeve balun.

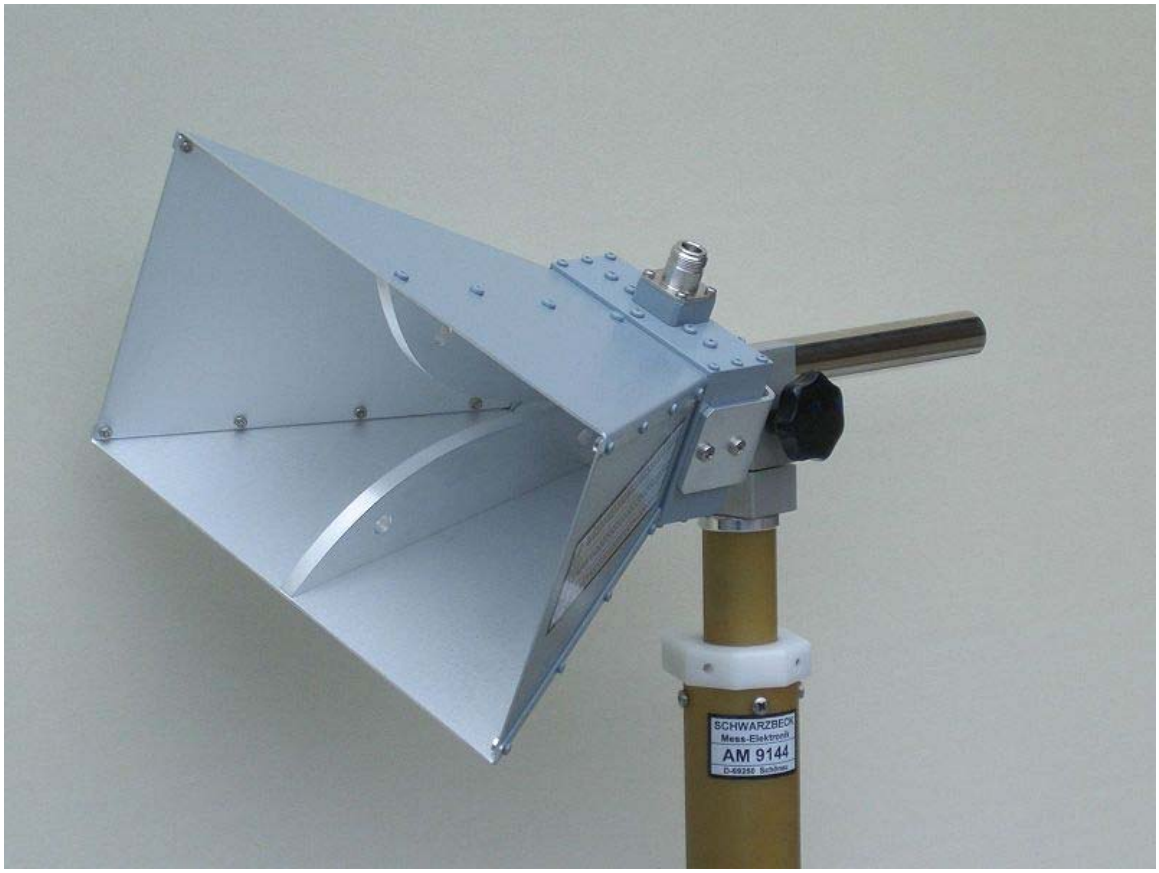
At VHF frequencies, a sleeve balun can also be built to remove feeder radiation.

- Another narrow band design is to use a λ/4 length of metal pipe. The coaxial cable is placed inside the pipe; at one end the braid is wired to the pipe while at the other end no connection is made to the pipe. The balanced end of this balun is at the end where the pipe is wired to the braid. The λ/4 conductor acts as a transformer converting the infinite impedance at the unconnected end into a zero impedance at the end connected to the braid. Hence any current entering the balun through the connection, which goes to the braid at the end with the connection to

the pipe, will flow into the pipe. This balun design is impractical for low frequencies because of the long length of pipe that will be needed.

Chapter 2

Horn Antenna



Pyramidal microwave horn antenna, with a bandwidth of 0.8 to 18 GHz. A coaxial cable feedline attaches to the connector visible at top. This type is called a ridged horn; the curving fins visible inside the mouth of the horn increase the antenna's bandwidth.

A **horn antenna** or **microwave horn** is an antenna that consists of a flaring metal waveguide shaped like a horn to direct the radio waves. Horns are widely used as antennas at UHF and microwave frequencies, above 300 MHz. They are used as feeders

(called feed horns) for larger antenna structures such as parabolic antennas, as standard calibration antennas to measure the gain of other antennas, and as directive antennas for such devices as radar guns, automatic door openers, and microwave radiometers. Their advantages are moderate directivity (gain), low SWR, broad bandwidth, and simple construction and adjustment.

One of the first horn antennas was constructed in 1897 by Indian radio researcher Jagadish Chandra Bose in his pioneering experiments with microwaves. In the 1930s the first experimental research (Southworth and Barrow, 1936) and theoretical analysis (Barrow and Chu, 1939) of horns as antennas was done. The development of radar in World War 2 stimulated horn research. The corrugated horn proposed by Kay in 1962 has become widely used as a feed horn for microwave antennas such as satellite dishes and radio telescopes.

An advantage of horn antennas is that since they don't have any resonant elements, they can operate over a wide range of frequencies, a wide bandwidth. The useable bandwidth of horn antennas is typically of the order of 10:1, and can be up to 20:1 (for example allowing it to operate from 1 GHz to 20 GHz). The input impedance is slowly-varying over this wide frequency range, allowing low VSWR over the bandwidth. The gain of horn antennas ranges up to 25 dBi, with 10 - 20 dBi being typical.

Description

A horn antenna is used to transmit radio waves from a waveguide (a metal pipe used to carry radio waves) out into space, or collect radio waves into a waveguide for reception. It typically consists of a short length of rectangular or cylindrical metal tube (the waveguide), closed at one end, flaring into an open-ended conical or pyramidal shaped horn on the other end. The radio waves are usually introduced into the waveguide by a coaxial cable attached to the side, with the central conductor projecting into the waveguide. The waves then radiate out the horn end in a narrow beam. However in some equipment the radio waves are conducted from the transmitter or to the receiver by a waveguide, and in this case the horn is just attached to the end of the waveguide.

How it works

Corrugated conical horn antenna used as a feed horn on a Hughes Direcway home satellite dish.

A horn antenna serves the same function for electromagnetic waves that an acoustical horn does for sound waves in a musical instrument such as a trumpet; it provides a gradual transition structure to match the impedance of a tube to the impedance of free space, enabling the waves from the tube to radiate efficiently into space.

If a simple open-ended waveguide were to be used as an antenna, without the horn, the sudden end of the conductive walls causes an abrupt impedance change at the aperture, from the characteristic impedance of the waveguide to the impedance of free space, 377 ohms. When radio waves travelling through the waveguide hit the opening, it acts as a bottleneck, reflecting most of the wave energy back down the guide toward the source, so only part of the power is radiated. It acts similarly to an open-circuited transmission line, or to a boundary between optical mediums with a high and low index of refraction, like a glass surface. The reflected waves cause standing waves in the waveguide, increasing the VSWR, wasting energy and possibly overheating the transmitter. In addition, the small aperture of the waveguide (around one wavelength) causes severe diffraction of the waves issuing from it, resulting in a wide radiation pattern without much directivity.

To improve these poor characteristics, the ends of the waveguide are flared out to form a horn. The taper of the horn changes the impedance gradually along the horn's length. This acts like an impedance matching transformer, allowing most of the wave energy to radiate out the end of the horn into space, with minimal reflection. The taper functions similarly to a tapered transmission line, or an optical medium with a smoothly-varying refractive index. In addition, the wide aperture of the horn projects the waves in a narrow beam

The horn shape that gives minimum reflected power is an exponential taper. Exponential horns are used in special applications that require minimum signal loss, such as satellite antennas and radio telescopes. However conical and pyramidal horns are most widely used, because they have straight sides and are easier to fabricate.

Radiation pattern

The waves travel down a horn as spherical wavefronts, with their origin at the apex of the horn. The pattern of electric and magnetic fields at the aperture plane of the horn, which determines the radiation pattern, is a scaled-up reproduction of the fields in the waveguide. However, because the wavefronts are spherical, the phase increases smoothly from the center of the aperture plane to the edges, because of the difference in length of the center point and the edge points from the apex point. The difference in phase between the center point and the edges is called the *phase error*. This phase error, which increases with the flare angle, reduces the gain and increases the beamwidth, giving horns wider beamwidths than plane-wave antennas such as parabolic dishes.

At the flare angle, the radiation of the beam lobe is down about -20 dB from its maximum value.

The increasing phase error limits the aperture size of practical horns to about 15 wavelengths; larger apertures would require impractically long horns. This limits the gain of practical horns to about 1000 (30 dB) and the corresponding minimum beamwidth to about 5 - 10°.

Optimum horn

Large pyramidal horn used in 1951 to detect the 21 cm (1.43 GHz) radiation from hydrogen gas in the Milky Way galaxy.

For a given frequency and horn length, there is some flare angle that gives minimum reflection and maximum gain. The reflections in straight-sided horns come from the two locations along the wave path where the impedance changes abruptly; the mouth or aperture of the horn, and the throat where the sides begin to flare out. The amount of reflection at these two sites varies with the *flare angle* of the horn (the angle the sides make with the axis). In narrow horns with small flare angles most of the reflection occurs at the mouth of the horn. The gain of the antenna is low because the small mouth approximates an open-ended waveguide. As the angle is increased, the reflection at the mouth decreases rapidly and the antenna's gain increases. In contrast, in wide horns with

flare angles approaching 90° most of the reflection is at the throat. The horn's gain is again low because the throat approximates an open-ended waveguide. As the angle is decreased, the amount of reflection at this site drops, and the horn's gain again increases.

This discussion shows that there is some flare angle which gives maximum gain and minimum reflection. This is called the *optimum horn*. Most practical horn antennas are designed as optimum horns. In a pyramidal horn, the dimensions that give an optimum horn are:

$$a_E = \sqrt{3\lambda L_E} \qquad a_H = \sqrt{2\lambda L_H}$$

For a conical horn, the dimensions that give an optimum horn are:

$$d = \sqrt{3\lambda L}$$

where

a_E is the width of the aperture in the E-field direction
a_H is the width of the aperture in the H-field direction
L_E is the slant length of the side in the E-field direction
L_H is the slant length of the side in the H-field direction.
d is the diameter of the cylindrical horn aperture
L is the slant length of the cone from the apex.
λ is the wavelength

An optimum horn does not give maximum gain for a given *aperture size*; this is achieved by a very long horn. It gives the maximum gain for a given horn *length*. Tables showing dimensions for optimum horns for various frequencies are given in microwave handbooks.

Gain

Horns have very little loss, so the directivity of a horn is roughly equal to its gain. The gain G of a pyramidal horn antenna (the ratio of the radiated power intensity along its beam axis to the intensity of an isotropic antenna with the same input power) is:

$$G = \frac{4\pi A}{\lambda^2} e_A$$

For conical horns, the gain is:

$$G = \left(\frac{\pi d}{\lambda}\right)^2 e_A$$

where

A is the area of the aperture,
d is the aperture diameter of a conical horn
λ is the wavelength,
e_A is a dimensionless parameter called the *aperture efficiency*,

The aperture efficiency ranges from 0.4 to 0.8 in practical horn antennas. For optimum pyramidal horns, $e_A = 0.511$., while for optimum conical horns $e_A = 0.522$. So an approximate figure of 0.5 is often used. The aperture efficiency increases with the length of the horn, and for aperture-limited horns is approximately unity.

Types of horn antennas

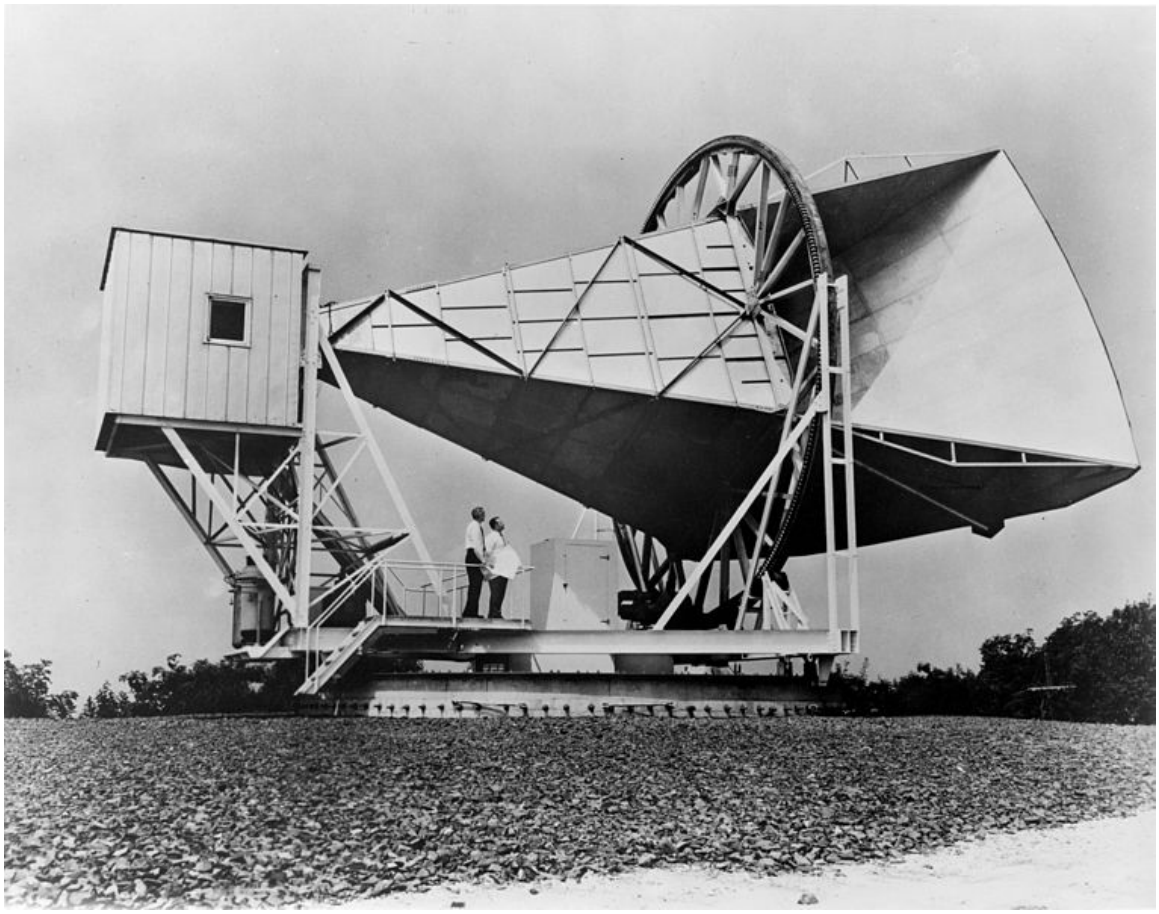
Aperture-limited corrugated horn, used as a feed horn in a radio telescope for millimeter waves.

These are the common types of horn antenna. Horns can have different flare angles as well as different expansion curves (elliptic, hyperbolic, etc.) in the E-field and H-field directions, making possible a wide variety of different beam profiles.

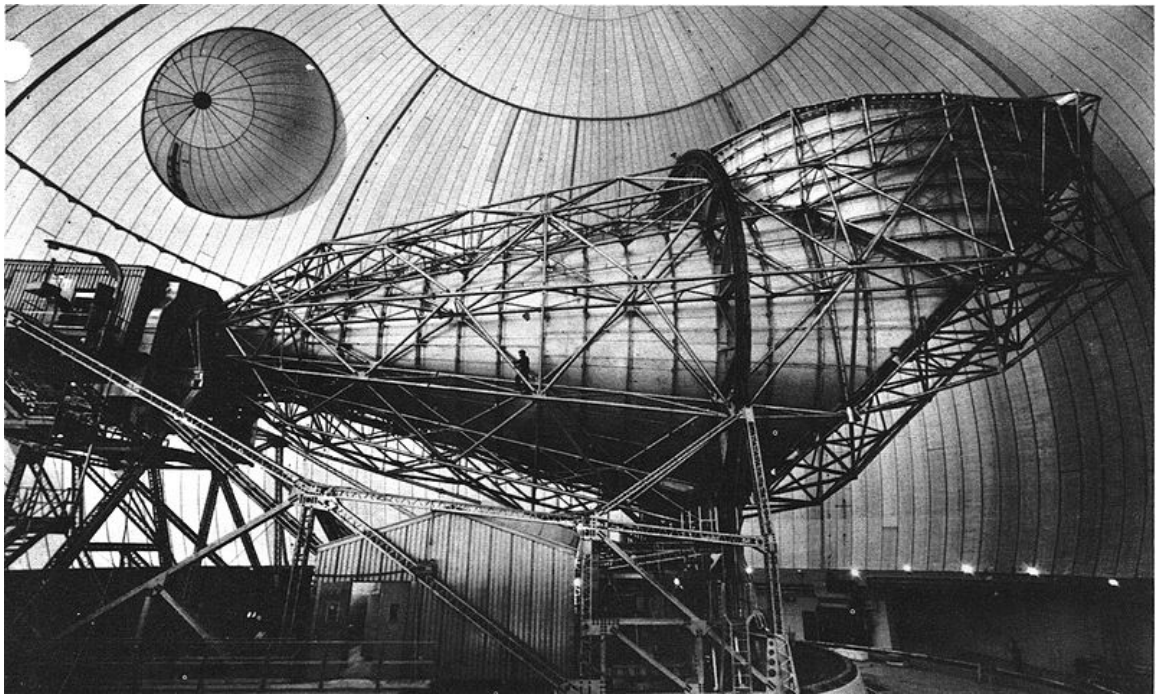
- Pyramidal horn - a horn antenna with the horn in the shape of a four-sided pyramid, with a rectangular cross section. They are the most widely used type, used with rectangular waveguides, and radiate linearly polarized radio waves.
- Sectoral horn - A pyramidal horn with only one pair of sides flared and the other pair parallel. It produces a fan-shaped beam, which is narrow in the plane of the flared sides, but wide in the plane of the narrow sides.
    - E-plane horn - A sectoral horn flared in the direction of the electric or E-field in the waveguide.
    - H-plane horn - A sectoral horn flared in the direction of the magnetic or H-field in the waveguide.
- Conical horn - A horn in the shape of a cone, with a circular cross section. They are used with cylindrical waveguides.
- Corrugated horn - A horn with parallel slots or grooves, small compared with a wavelength, covering the inside surface of the horn, transverse to the axis. Corrugated horns have wider bandwidth and smaller sidelobes and cross-

polarization, and are widely used as feed horns for satellite dishes and radio telescopes.

- Ridged horn - A pyramidal horn with ridges or fins attached to the inside of the horn, extending down the center of the sides. The fins lower the cutoff frequency, increasing the antenna's bandwidth.
- Septum horn - A horn which is divided into several subhorns by metal partitions (septums) inside, attached to opposite walls.
- Aperture-limited horn - a long narrow horn, long enough so the phase error is a fraction of a wavelength, so it essentially radiates a plane wave. It has an aperture efficiency of 1.0 so it gives the maximum gain and minimum beamwidth for a given aperture size. The gain is not affected by the length but only limited by diffraction at the aperture. Used as feed horns in radio telescopes and other high-resolution antennas.



50 ft. Holmdel horn antenna at Bell labs in Holmdel, New Jersey, USA, with which Arno Penzias and Robert Wilson discovered cosmic microwave background radiation in 1964.



Large 177 ft. Hogg horn antenna at AT&T satellite communications facility in Andover, Maine, USA, used in 1960s to communicate with the first direct relay communications satellite, Telstar.

Hogg microwave relay antennas on roof of AT&T telephone switching center, Seattle, Washington, USA
Hogg antennas

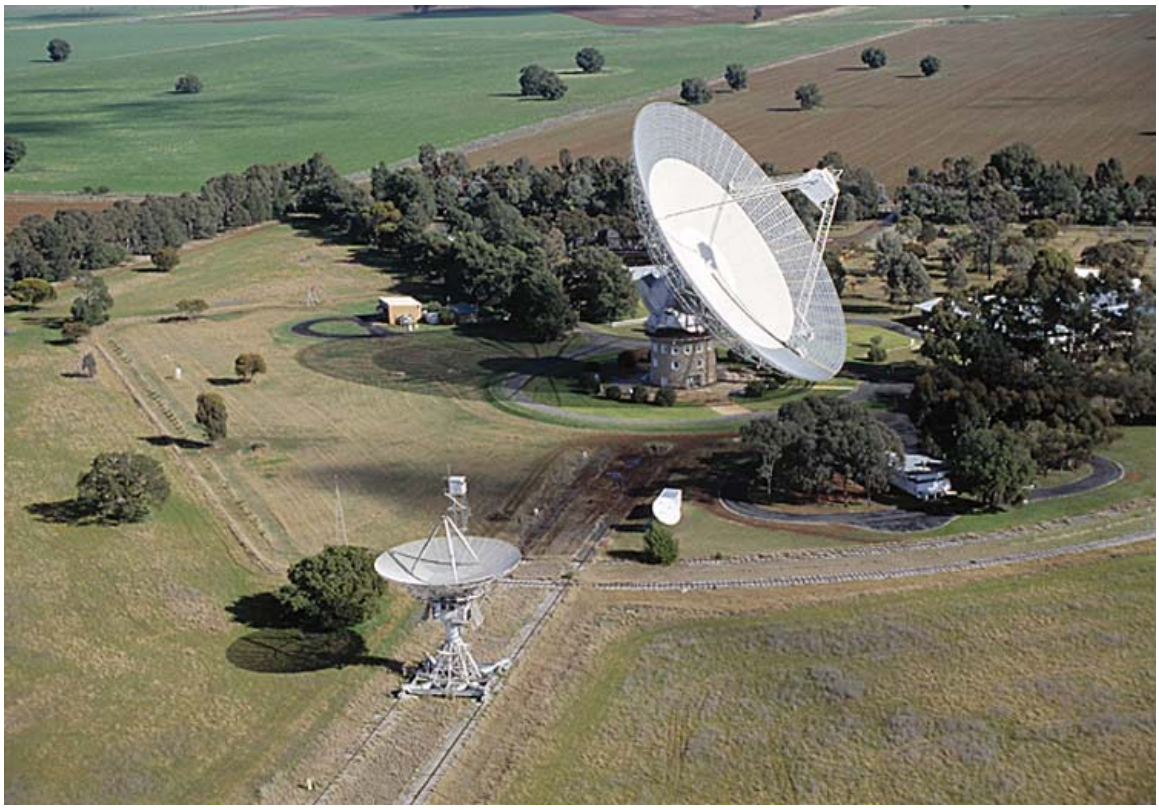
Hogg horn antenna

A type of antenna that combines a horn with a parabolic reflector is the Hogg antenna, invented by D. C. Hogg at Bell labs around 1960. It consisted of a horn antenna with a reflector mounted in the mouth of the horn at a 45 degree angle so the radiated beam is at right angles to the horn axis. The reflector is a segment of a parabolic reflector, so the device is equivalent to a parabolic antenna fed off-axis. The advantage of this design over a standard parabolic antenna is that the horn shields the antenna from radiation coming from angles outside the main beam axis, so its radiation pattern has very small sidelobes.

Also, the aperture isn't partially obstructed by the feed and its supports, as with ordinary front-fed parabolic dishes. The disadvantage is that it is far larger and heavier for a given aperture area than a parabolic dish, and must be mounted on a cumbersome turntable to be fully steerable. This design was used for a few radio telescopes and communication satellite ground antennas during the 1960s. Its largest use, however, was as fixed antennas for microwave relay links in the AT&T Long Lines microwave network. Probably the most photographed and well-known example is the 15 meter (50 foot) long Holmdel Horn Antenna at Bell Labs in Holmdel, New Jersey, with which Arno Penzias and Robert Wilson discovered cosmic microwave background radiation in 1965, for which they won the 1978 Nobel Prize in Physics.

Chapter 3

Radio Telescope



The 64 meter radio telescope at Parkes Observatory

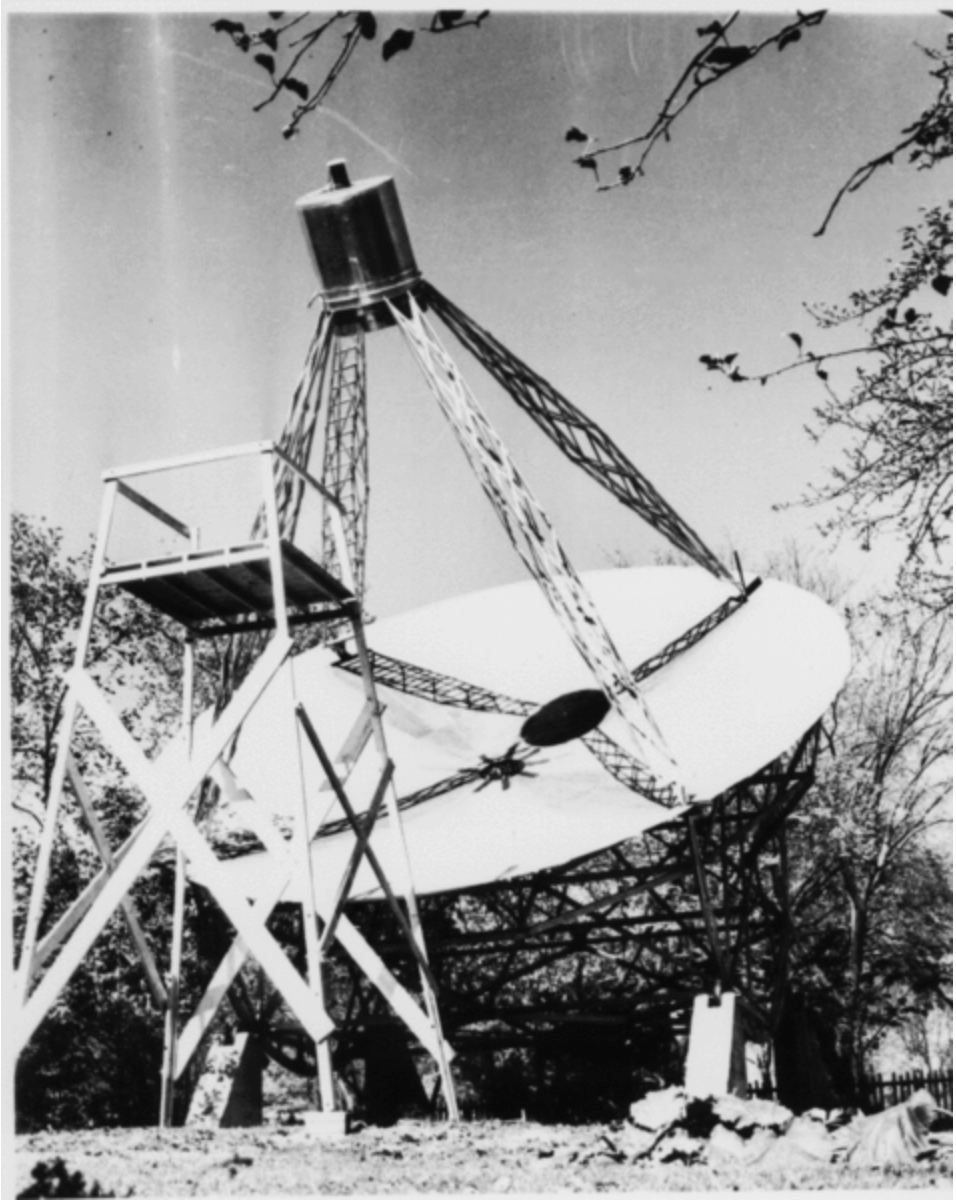
A **radio telescope** is a form of directional radio antenna used in radio astronomy. The same types of antennas are also used in tracking and collecting data from satellites and space probes. In their astronomical role they differ from optical telescopes in that they operate in the radio frequency portion of the electromagnetic spectrum where they can detect and collect data on radio sources. Radio telescopes are typically large parabolic ("dish") antennas used singly or in an array. Radio observatories are preferentially located far from major centers of population to avoid electromagnetic interference (EMI) from radio, TV, radar, and other EMI emitting devices. This is similar to the locating of optical

telescopes to avoid light pollution, with the difference being that radio observatories are often placed in valleys to further shield them from EMI as opposed to clear air mountain tops for optical observatories.

Early radio telescopes

Full-size replica of the first radio telescope, Jansky's dipole array now at the US National Radio Astronomy Observatory

The first radio antenna used to identify an astronomical radio source was one built by Karl Guthe Jansky, an engineer with Bell Telephone Laboratories, in 1931. Jansky was assigned the job of identifying sources of static that might interfere with radio telephone service. Jansky's antenna was an array of dipoles and reflectors designed to receive short wave radio signals at a frequency of 20.5 MHz (wavelength about 14.6 metres). It was mounted on a turntable that allowed it to rotate in any direction, earning it the name "Jansky's merry-go-round". It had a diameter of approximately 100 ft (30 m). and stood 20 ft (6 m). tall. By rotating the antenna on a set of four Ford Model-T tires, the direction of the received interfering radio source (static) could be pinpointed. A small shed to the side of the antenna housed an analog pen-and-paper recording system. After recording signals from all directions for several months, Jansky eventually categorized them into three types of static: nearby thunderstorms, distant thunderstorms, and a faint steady hiss of unknown origin. Jansky finally determined that the "faint hiss" repeated on a cycle of 23 hours and 56 minutes. This period is the length of an astronomical sidereal day, the time it takes any "fixed" object located on the celestial sphere to come back to the same location in the sky. Thus Jansky suspected that the hiss originated well beyond the Earth's atmosphere, and by comparing his observations with optical astronomical maps, Jansky concluded that the radiation was coming from the Milky Way Galaxy and was strongest in the direction of the center of the galaxy, in the constellation of Sagittarius.



Reber's first "dish" radio telescope - Wheaton, IL 1937

An amateur radio operator, Grote Reber, was one of the pioneers of what became known as radio astronomy when he built the first parabolic "dish" radio telescope (9 metres (30 ft) in diameter) in his back yard in Illinois in 1937. He was instrumental in repeating Karl Guthe Jansky's pioneering but somewhat simple work at higher frequencies, and he went on to conduct the first sky survey at very high radio frequencies. The rapid development of radar technology during World War II was easily translated into radio astronomy technology after the war, and the field of radio astronomy began to blossom.

Types

A *cylindrical paraboloid* antenna.

The range of frequencies in the electromagnetic spectrum that makes up the radio spectrum is very large. This means that the types of antennas that are used as radio telescopes vary widely in design, size, and configuration. At wavelengths of 30 meters to 3 meters (10 MHz - 100 MHz), they are generally either directional antenna arrays similar to "TV antennas" or large stationary reflectors with moveable focal points. Since the wavelengths being observed with these types of antennas are so long, the "reflector" surfaces can be constructed from coarse wire mesh such as chicken wire. At shorter wavelengths “dish” style radio telescopes predominate. The angular resolution of a dish style antenna is determined by the diameter of the dish expressed as a number of

wavelengths of the electromagnetic radiation being observed. This dictates the dish size a radio telescope needs for a useful resolution. Radio telescopes that operate at wavelengths of 3 meters to 30 cm (100 MHz to 1 GHz) are usually well over 100 meters in diameter. Telescopes working at wavelengths shorter than 30 cm (above 1 GHz) range in size from 3 to 90 meters in diameter.

Big dishes

RT-70 planetary radar

World's largest single-aperture radio telescope at Arecibo Observatory in Puerto Rico

ALMA construction site

The world's largest filled-aperture telescope (i.e., a full dish) is the Arecibo radio telescope located in Arecibo, Puerto Rico, whose 305 m (1,001 ft) dish is fixed in the ground. It was designed by engineer Bill Gordon (d. 2010). The suspension system was designed by George and Helias Doundoulakis, for which Helias Doundoulakis received a patent along with assignee William J. Casey, ex-Central Intelligence Agency Director under President Ronald Reagan. The antenna beam is steerable (by means of a moving receiver) within about 20° of the zenith. The largest individual radio telescope of any kind is the RATAN-600 located near Nizhny Arkhyz, Russia, which consists of a 576-meter circle of rectangular radio reflectors, each of which can be pointed towards a central conical receiver.

The largest radio telescope in Europe is the 100-meter diameter antenna in Effelsberg, Germany, which also was the world's largest fully-steerable telescope for 30 years until the slightly larger Green Bank Telescope was opened in West Virginia, United States, in 2000. The third-largest fully-steerable radio telescope is the 76-metre Lovell Telescope at Jodrell Bank Observatory in Cheshire, England. The fourth-largest fully-steerable radio telescopes is four the 70-metre radio telescopes: RT-70, Goldstone. Feature of these telescopes is that it is the world's largest planetary radars (except Suffa RT-70).

A typical size of the single antenna of a radio telescope is 25 meters. Dozens of radio telescopes with comparable sizes are operated in radio observatories all over the world.

China officially started construction of the world's largest single-aperture radio telescope in 2009, the FAST. The FAST, with a dish area as large as 30 football fields, will stand in a region of typical Karst depressions in Guizhou, and will be finished by 2013.

Radio interferometry

The Very Large Array, an interferometric array formed from many smaller telescopes, like many larger radio telescopes.

One of the most notable developments came in 1946 with the introduction of the technique called astronomical interferometry. Astronomical radio interferometers usually consist either of arrays of parabolic dishes (e.g., the One-Mile Telescope), arrays of one-dimensional antennas (e.g., the Molonglo Observatory Synthesis Telescope) or two-dimensional arrays of omni-directional dipoles (e.g., Tony Hewish's Pulsar Array). All of the telescopes in the array are widely separated and are usually connected together using coaxial cable, waveguide, optical fiber, or other type of transmission line. Recent advances in the stability of electronic oscillators also now permit interferometry to be carried out by independent recording of the signals at the various antennas, and then later correlating the recordings at some central processing facility. This process is known as VLBI (Very Long Baseline Interferometry). Interferometry does increase the total signal collected, but its primary purpose is to vastly increase the resolution through a process called Aperture synthesis. This technique works by superposing (interfering) the signal waves from the different telescopes on the principle that waves that coincide with the

same phase will add to each other while two waves that have opposite phases will cancel each other out. This creates a combined telescope that is equivalent in resolution (though not in sensitivity) to a single antenna whose diameter is equal to the spacing of the antennas furthest apart in the array. A high quality image requires a large number of different separations between telescopes. Projected separation between any two telescopes, as seen from the radio source, is called a baseline. For example, the Very Large Array (VLA) near Socorro, New Mexico has 27 telescopes with 351 independent baselines at once, which achieves a resolution of 0.2 arc seconds at 3 cm wavelengths). Martin Ryle's group in Cambridge obtained a Nobel Prize for interferometry and aperture synthesis. The Lloyd's mirror interferometer was also developed independently in 1946 by Joseph Pawsey's group at the University of Sydney. In the early 1950s the Cambridge Interferometer mapped the radio sky to produce the famous 2C and 3C surveys of radio sources. The largest existing physically-connected radio telescope array is the Giant Metrewave Radio Telescope, located in Pune, India. A larger array, LOFAR (the 'LOw Frequency ARray') is currently being constructed in western Europe, consisting of 25 000 small antennas over an area several hundreds of kilometres in diameter. VLBI systems using post-observation processing have been constructed with antennas thousands of miles apart. Radio interferometers have also been used to obtain detailed images of the anisotropies and the polarization of the Cosmic Microwave Background, like the CBI interferometer in 2004.

Astronomical observations

Many astronomical objects are not only observable in visible light but also emit radiation at radio wavelengths. Besides observing energetic objects such as pulsars and quasars, radio telescopes are able to "image" most astronomical objects such as galaxies, nebulae, and even radio emissions from planets.

Chapter 4

Parabolic Antenna

A parabolic antenna at Erdfunkstelle Raisting, the biggest facility for satellite communication in the world, based in Raisting, Bavaria, Germany. It has a Cassegrain type feed.

A **parabolic antenna** is an antenna that uses a parabolic reflector, a surface with the cross-sectional shape of a parabola, to direct the radio waves. The most common form is shaped like a dish and is popularly called a **dish antenna** or **parabolic dish**. The main advantage of a parabolic antenna is that it is highly directive; it functions analogously to a searchlight or flashlight reflector to direct the radio waves in a narrow beam, or receive radio waves from one particular direction only. Parabolic antennas have some of the highest gains, that is they can produce the narrowest beam width angles, of any antenna type. They are used as high-gain antennas for point-to-point radio, television and data communications, and also for radiolocation (radar), on the UHF and microwave (SHF) parts of the radio spectrum. The relatively short wavelength of electromagnetic radiation at these frequencies allows reasonably sized reflectors to exhibit the desired highly directional response.

With the advent of TVRO and DBS satellite television dishes, parabolic antennas have become a ubiquitous feature of the modern landscape. They are also widely used for terrestrial microwave relay links, ground based and airborne radar antennas, wireless WAN/LAN links, satellite and spacecraft communication antennas, and radio telescopes.

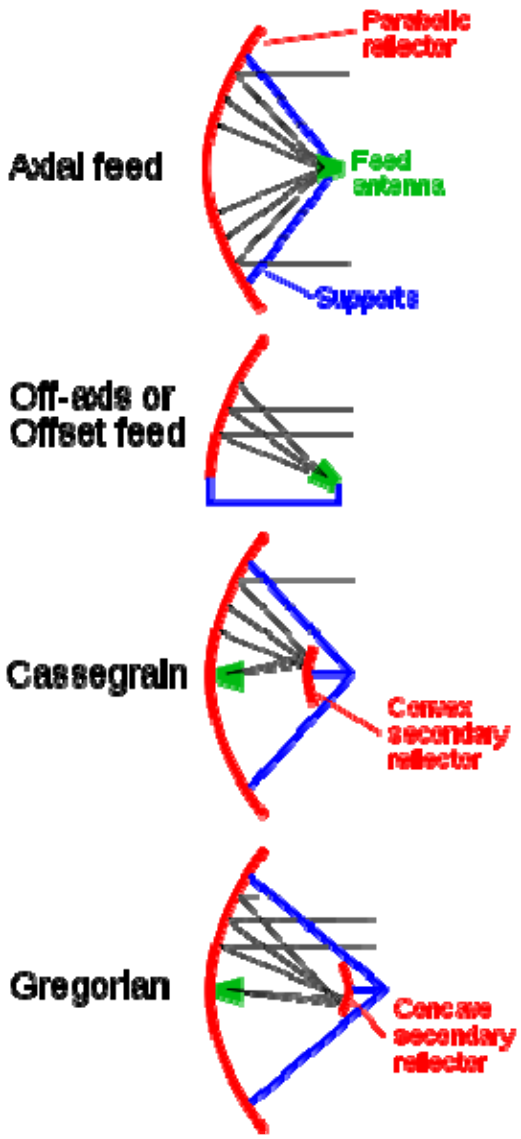
History

The idea of using parabolic reflectors for radio antennas was taken from optics, where the power of a parabolic mirror to focus light into a beam has been known since classical antiquity. The designs of some specific types of parabolic antenna, such as the Cassegrain and Gregorian, come from similarly named analogous types of reflecting telescope, which were invented by astronomers during the 15th century.

German physicist Heinrich Hertz constructed the world's first parabolic reflector antenna in 1888. The antenna was a cylindrical parabolic reflector made of zinc sheet metal supported by a wooden frame, and had a spark-gap excited dipole along the focal line. Its aperture was 1.2 meters wide, with a focal length of 0.12 meters, and was used at an operating frequency of about 450 MHz. With two such antennas, one used for transmitting and the other for receiving, Hertz demonstrated the existence radio waves which had been predicted by James Clerk Maxwell some 22 years earlier.

The first parabolic antenna used for satellite communications was established in 1962 at Goonhilly in Cornwall, England, UK and was intended to communicate with Telstar.

Types



Main types of parabolic antenna feeds.

Parabolic antennas are distinguished by their shapes:

- *Paraboloidal* or *dish* - The reflector is shaped like a paraboloid. This is the most common type. It radiates a narrow pencil-shaped beam along the axis of the dish.
    - *Shrouded dish* - Sometimes a cylindrical metal shield is attached to the rim of the dish. The shroud shields the antenna from radiation from angles outside the main beam axis, reducing the sidelobes. It is sometimes used to prevent interference in terrestrial microwave links, where several antennas using the same frequency are located close together. The shroud is coated

inside with microwave absorbent material. Shrouds can reduce back lobe radiation by 10 dB.

- *Cylindrical* - The reflector is curved in only one direction and flat in the other. The radio waves come to a focus not at a point but along a line. The feed is often a dipole antenna located along the focal line. It radiates a fan-shaped beam, narrow in the curved dimension, and wide in the uncurved dimension. The curved ends of the reflector are sometimes capped by flat plates, to prevent radiation out the ends, and this is called a *pillbox* antenna.
    - *"Orange peel"* - Another type is very long and narrow, shaped like the letter "C". This is called an *orange peel* design, and radiates an even wider fan beam. It is often used for radar antennas.

They are also classified by the type of *feed*; how the radio waves are supplied to the antenna:

- *Axial* or *front feed* - This is the most common type of feed, with the feed antenna located in front of the dish at the focus, on the beam axis. A disadvantage of this type is that the feed and its supports block some of the beam, which limits the aperture efficiency to only 55 - 60%.
- *Offset* or *off-axis feed* - The reflector is an asymmetrical segment of a paraboloid, so the focus, and the feed antenna, are located to one side of the dish. The purpose of this design is to move the feed structure out of the beam path, so it doesn't block the beam. It is widely used in home satellite television dishes, which are small enough that the feed structure would otherwise block a significant percentage of the signal.
- *Cassegrain* - In a Cassegrain antenna the feed is located on or behind the dish, and radiates forward, illuminating a convex hyperboloidal secondary reflector at the focus of the dish. The radio waves from the feed reflect back off the secondary reflector to the dish, which forms the main beam. An advantage of this configuration is that the feed, with its waveguides and "front end" electronics does not have to be suspended in front of the dish, so it is used for antennas with complicated or bulky feeds, such as large satellite communication antennas and radio telescopes. Aperture efficiency is on the order of 65 - 70%
- *Gregorian* - Similar to the Cassegrain design except that the secondary reflector is concave, (ellipsoidal) in shape. Aperture efficiency over 70% can be achieved.

Design

The operating principle of a parabolic antenna is that a point source of radio waves at the focal point in front of a parabolic reflector will be reflected into a collimated plane wave beam along the axis of the reflector. Conversely, an incoming plane wave parallel to the axis will be focused to a point at the focal point.

A typical parabolic antenna consists of a parabolic reflector with a small feed antenna at its focus, pointed back toward the reflector. The reflector is a metallic surface formed into

a paraboloid of revolution and usually truncated in a circular rim that forms the diameter of the antenna.

The reflector dish can be of sheet metal, metal screen, or wire grill construction, and it can be either circular or various other shapes to create different beam shapes. A mesh screen reflects radio waves as well as a solid metal surface as long as the holes are smaller than 1/10 of a wavelength, so screen reflectors are often used to reduce weight and wind loads on the dish. To achieve the maximum gain, it is necessary that the shape of the dish be accurate within a small fraction of a wavelength, to ensure the waves from different parts of the antenna arrive in phase. Large dishes often require a truss structure behind them to provide the required stiffness.

The feed antenna at the reflector's focus is typically a low-gain type such as a half-wave dipole or more ofteh a small horn antenna called a feed horn. In more complex designs, such as the Cassegrain and Gregorian, a secondary reflector is used to direct the energy into the parabolic reflector from a feed antenna located away from the primary focal point. The feed antenna is connected to the associated radio-frequency (RF) transmitting or receiving equipment by means of a coaxial cable transmission line or waveguide.

Feed pattern

The radiation pattern of the feed antenna has a strong influence on the *aperture efficiency*, which determines the antenna gain. Radiation from the feed that falls outside the edge of the dish is called "*spillover*" and is wasted, reducing the gain and increasing the backlobes, possibly causing interference or (in receiving antennas) increasing susceptibility to ground noise. However, maximum gain is only achieved when the dish is uniformly "illuminated" with a constant field strength to its edges. So the ideal radiation pattern of a feed antenna would be a constant field strength throughout the solid angle of the dish, dropping abruptly to zero at the edges. However, practical feed antennas have radiation patterns that drop off gradually at the edges, so the feed antenna is a compromise between acceptably low spillover and adequate illumination.

Gain

The directive qualities of an antenna are measured by a dimensionless parameter called its gain, which is the ratio of the power received by the antenna from a source along its beam axis to the power received by a hypothetical isotropic antenna. The gain of a parabolic antenna is:

$$G = \frac{4\pi A}{\lambda^2} e_A = \frac{\pi^2 d^2}{\lambda^2} e_A$$

where:

- A is the area of the antenna aperture, that is, the mouth of the parabolic reflector
- d is the diameter of the parabolic reflector

- λ is the wavelength of the radio waves.
- e_A is a dimensionless parameter called the *aperture efficiency*. The aperture efficiency of typical parabolic antennas is 0.55 to 0.70.

The largest dish antenna in the world, the radio telescope at Arecibo Observatory, Puerto Rico, 305 meters (1000 feet) in diameter. It has a gain of 10 million, or 70 dBi, at 2.38 GHz.

It can be seen that, as with any *aperture antenna*, the larger the aperture is, compared to the wavelength, the higher the gain. The gain increases with the square of the ratio of aperture width to wavelength, so large parabolic antennas, such as those used for spacecraft communication and radio telescopes, can have extremely high gain. Applying the above formula to the 25-meter-diameter antennas used by the VLA and VLBA radio telescopes at a wavelength of 21 cm (1.42 GHz, a common radio astronomy frequency) yields an approximate maximum gain of 140,000 times or about 50 dBi (decibels above the isotropic level).

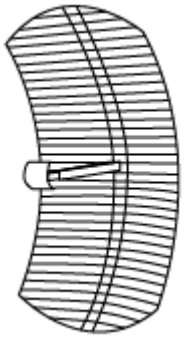
As of 2009, the largest "dish" antenna in the world is the Arecibo Observatory's radio telescope, at Arecibo, Puerto Rico, which has a diameter of 1,000 ft. (305 m). It has a gain of about 10 million, or 70 dBi, at 2.38 GHz. The dish is built into a valley in the landscape, so it is not steerable. To steer the beam to different points in the sky, the feed antenna is moved. For this reason, the dish actually has a spherical rather than a parabolic

shape. A spherical reflector does not have a single focal "point", however, the Arecibo antenna is a three-reflector variety of Gregorian telescope, and uses its secondary and tertiary reflectors to focus the radio waves to a single point.

Aperture efficiency e_A is a catchall variable which accounts for various losses that reduce the gain of the antenna from the maximum that could be achieved with the given aperture. The major factors reducing the aperture efficiency in parabolic antennas are:.

- *Feed spillover* - Some of the radiation from the feed antenna falls outside the edge of the dish and so doesn't contribute to the main beam.
- *Feed illumination taper* - The maximum gain for any aperture antenna is only achieved when the intensity of the radiated beam is constant across the aperture area. However the radiation pattern from the feed antenna usually tapers off toward the outer part of the dish, so the outer parts of the dish are "illuminated" with a lower intensity of radiation. Even if the feed provided constant illumination across the angle subtended by the dish, the outer parts of the dish are farther away from the feed antenna than the inner parts, so the intensity would drop off with distance from the center. So the intensity of the beam radiated by a parabolic antenna is maximum at the center of the dish and falls off with distance from the axis, reducing the efficiency.
- *Aperture blockage* - In axis-fed parabolic dishes where the feed antenna is located in front of the dish in the beam path, the feed structure and its supports block some of the beam. In small dishes such as home satellite dishes, where the size of the feed structure is comparable with the size of the dish, this can seriously reduce the antenna gain. To prevent this problem these types of antennas often use an *off-axis* feed, where the feed antenna is located to one side, outside the beam area. The aperture efficiency for these types of antennas can reach 0.7 to 0.8.
- *Shape errors* - random surface errors in the shape of the reflector reduce efficiency. The loss is approximated by Ruze's Equation.

For theoretical considerations of mutual interference (at frequencies between 2 and c. 30 GHz - typically in the Fixed Satellite Service) where specific antenna performance has not been defined, a *reference antenna* based on Recommendation ITU-R S.465 is used to calculate the interference, which will include the likely sidelobes for off-axis effects.



Wire grid-type parabolic antenna (Wi-Fi / WLAN antenna) at 2.4 GHz).

Beamwidth

C-band Satellite dish, Bucharest, Romania

The angular width of the beam radiated by high-gain antennas is measured by the *half-power beam width* (HPBW), which is the angular separation between the points on the antenna radiation pattern at which the power drops to one-half (-3 dB) its maximum value. For parabolic antennas, the HPBW θ is given by:

$$\theta = k\lambda/d$$

where k is a factor which depends on the shape of the reflector and the feed illumination pattern. For a "typical" parabolic antenna $k = 70$ when θ is in degrees.

For a typical 2 meter satellite dish operating on C band (4 GHz), like the one shown at right, this formula gives a beamwidth of about 2.6°. For the Arecibo antenna at 2.4 GHz the beamwidth is 0.028°. It can be seen that parabolic antennas can produce very narrow

beams, and aiming them can be a problem. Some parabolic dishes are equipped with a boresight so they can be aimed accurately at the other antenna.

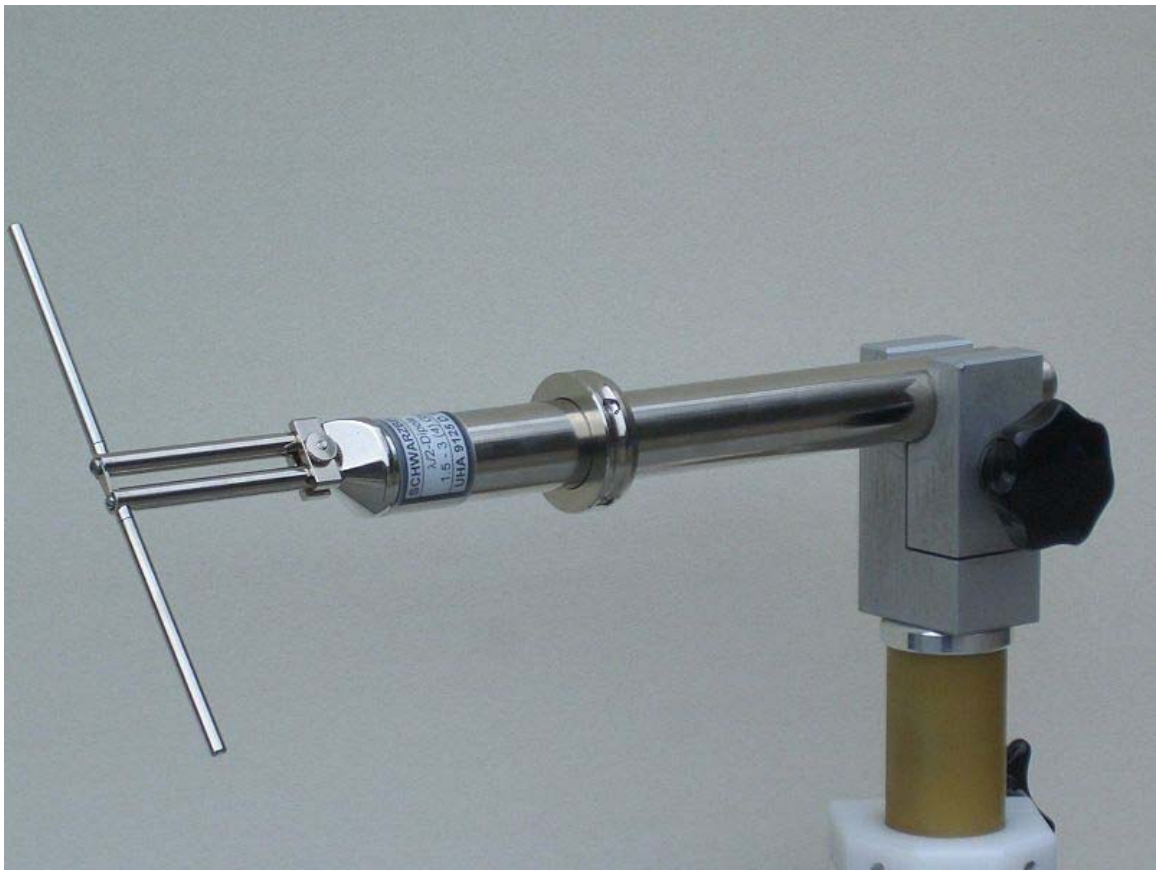
It can be seen there is an inverse relation between gain and beam width. By combining the beamwidth equation with the gain equation, the relation is:

$$G - \left(\frac{\pi k}{\theta}\right)^2 \epsilon_A$$

Chapter 5

Antenna (Radio)

Whip antenna on car



Half-wave dipole antenna

Large parabolic antenna for communicating with spacecraft

Rooftop directional antennas, typical for use at VHF and UHF frequencies

An **antenna** (or **aerial**) is an electrical device which couples radio waves in free space to an electrical current used by a radio receiver or transmitter. In reception, the antenna intercepts some of the power of an electromagnetic wave in order to produce a tiny voltage that the radio receiver can amplify. Alternatively, a radio transmitter will produce a large radio frequency current that may be applied to the terminals of the same antenna in order to convert it into an electromagnetic wave (radio wave) radiated into free space. Antennas are thus essential to the operation of all radio equipment, both transmitters and receivers. They are used in systems such as radio and television broadcasting, two-way radio, wireless LAN, mobile telephony, radar, and satellite communications.

Typically an antenna consists of an arrangement of metallic conductors (or "elements") with an electrical connection (often through a transmission line) to the receiver or transmitter. A current forced through such a conductor by a radio transmitter will create an alternating magnetic field according to Ampère's law. Or the alternating magnetic field due to a distant radio transmitter will induce a voltage at the antenna terminals, according to Faraday's law, which is connected to the input of a receiver. In the so-called far field, at a considerable distance away from the antenna, the oscillating magnetic field is coupled with a similarly oscillating electric field; together these define an electromagnetic wave which is capable of propagating great distances.

Light is one example of electromagnetic radiation, along with infrared and x-rays, while radio waves differ only in their much lower frequency (and much longer wavelength). Electronic circuits can operate at these lower frequencies, processing radio signals conducted through wires. But it is only through antennas that those radio frequency electrical signals are converted to (and from) propagating radio waves. Depending on the design of the antenna, radio waves can be sent toward and received from all directions ("omnidirectional"), whereas a directional or beam antenna is designed to operate in a particular direction.

The first antennas were built in 1888 by Heinrich Hertz (1857–1894) in his pioneering experiments to prove the existence of electromagnetic waves predicted by the theory of James Clerk Maxwell. Hertz placed dipole antennas at the focal point of parabolic reflectors for both transmitting and receiving. He published his work and installation drawings in *Annalen der Physik und Chemie* (vol. 36, 1889).

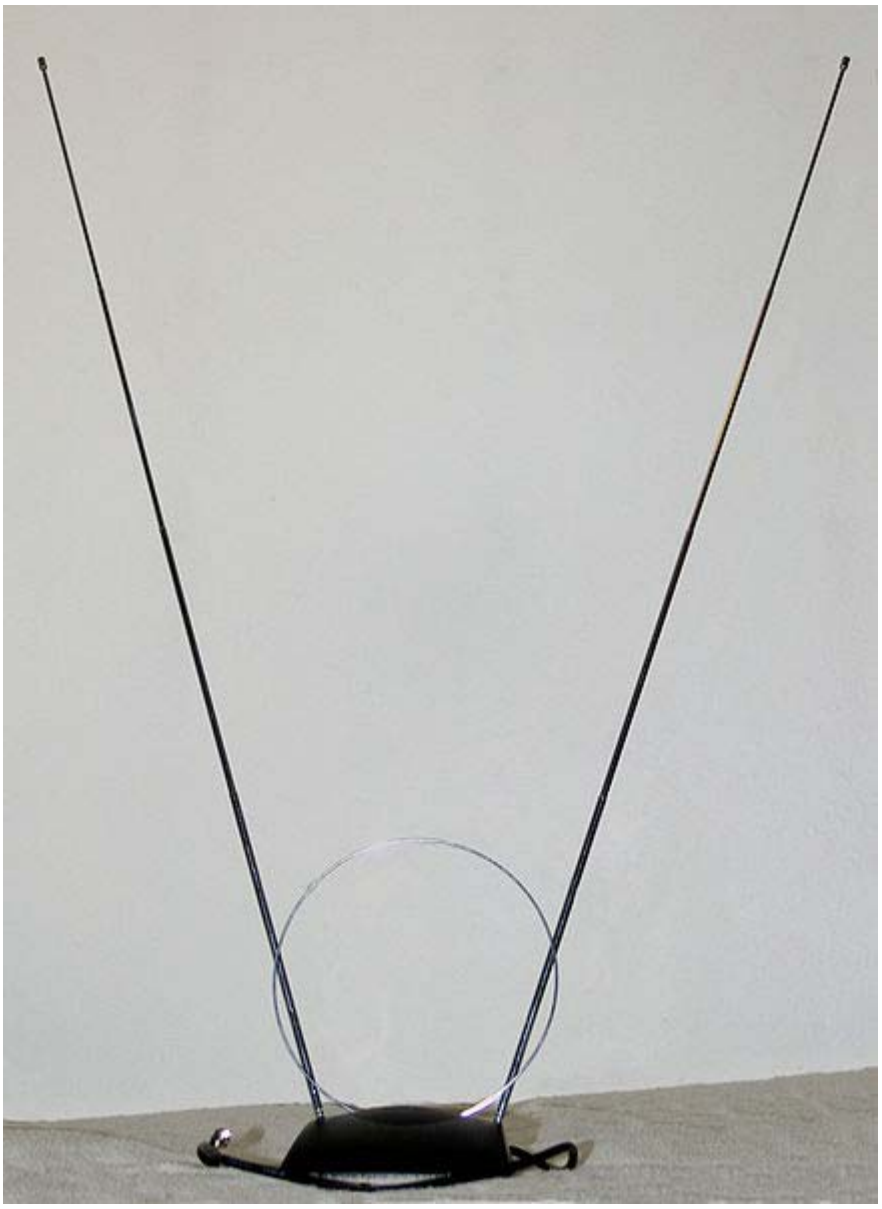
Terminology

The words *antenna* (plural: *antennas*) and *aerial* are used interchangeably; but usually a rigid metallic structure is termed an antenna and a wire format is called an aerial. In the United Kingdom and other British English speaking areas the term aerial is more common, even for rigid types. The noun *aerial* is occasionally written with a diaeresis mark—*aërial*—in recognition of the original spelling of the adjective *aërial* from which the noun is derived.

The origin of the word *antenna* relative to wireless apparatus is attributed to Guglielmo Marconi. In 1895, while testing early radio apparatuses in the Swiss Alps at Salvan, Switzerland in the Mont Blanc region, Marconi experimented with early wireless equipment. A 2.5 meter long pole, along which was carried a wire, was used as a radiating and receiving aerial element. In Italian a tent pole is known as *l'antenna centrale,* and the pole with a wire alongside it used as an aerial was simply called *l'antenna.* Until then wireless radiating transmitting and receiving elements were known simply as aerials or terminals. Marconi's use of the word *antenna* (Italian for *pole*) would become a popular term for what today is uniformly known as the *antenna*.

In common usage, the word *antenna* may refer broadly to an entire assembly including support structure, enclosure (if any), etc. in addition to the actual functional components.

Especially at microwave frequencies, a receiving antenna may include not only the actual electrical antenna but an integrated preamplifier and/or mixer.

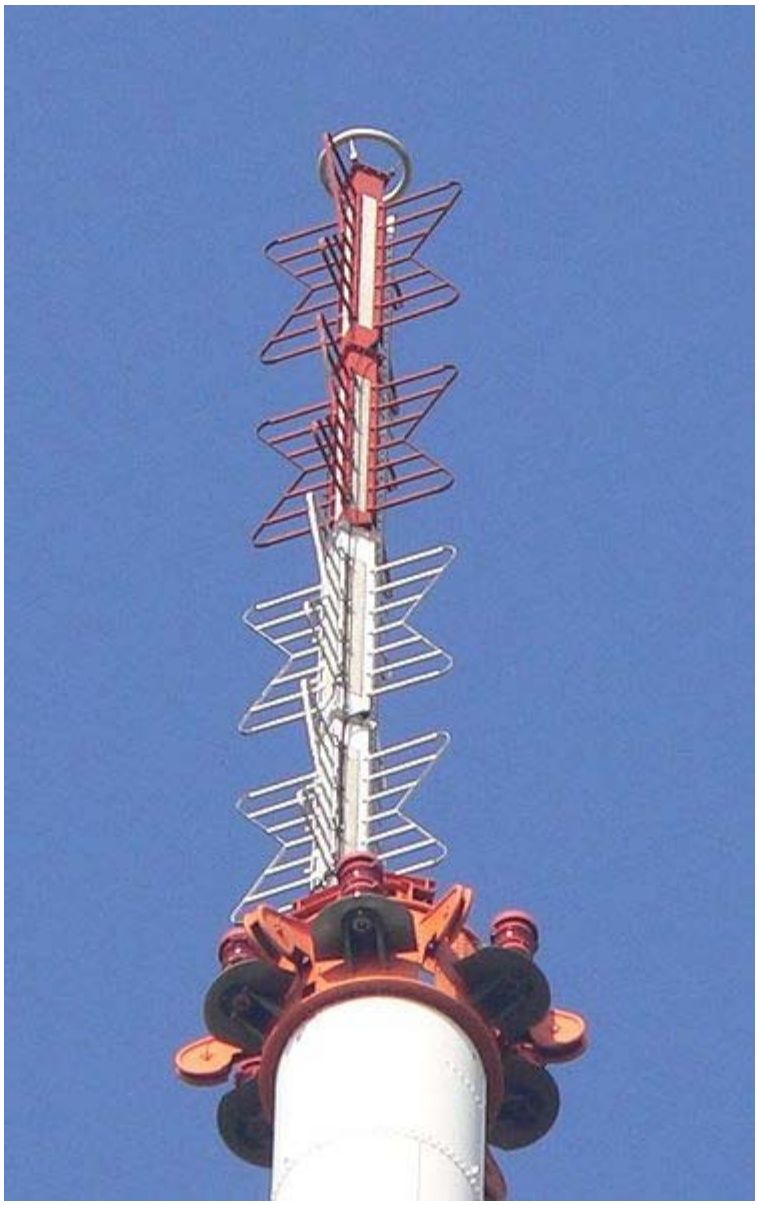


"Rabbit ears" dipole antenna for television reception

Cell phone base station antennas

Satellite link antenna used by Himalaya Television Nepal

Yagi antenna used for mobile military communications station, Dresden, Germany, 1955



"Super Turnstile" type transmitting antenna for VHF low band television broadcasting station, Germany.

Folded dipole antenna

Large Yagi antenna used by amateur radio hobbyist

A vertical mast radiator, Chapel Hill, North Carolina

Overview

Antennas are required by any radio receiver or transmitter in order to couple its electrical connection to the electromagnetic field. Radio waves are electromagnetic waves which carry signals through the air (or through space) at almost the speed of light with almost no transmission loss. Radio transmitters and receivers are used to convey signals (information) in systems including broadcast (audio) radio, television, mobile telephones, wi-fi (WLAN) data networks, trunk lines and point-to-point communications links (telephone, data networks), satellite links, many remote controlled devices such as garage door openers, and wireless remote sensors, among many others. Radio waves are also used directly for measurements in technologies including RADAR, GPS, and radio

astronomy. In each and every case, the transmitters and receivers involved require antennas, although these are sometimes hidden (such as the antenna inside an AM radio or inside a laptop computer equipped with wi-fi).

According to their applications and technology available, antennas generally fall in one of two catagories:

1. Omnidirectional or only weakly directional antennas which receive or radiate more or less in all directions. These are employed when the relative position of the other station is unknown or arbitrary. They are also used at lower frequencies where a directional antenna would be too large, or simply to cut costs in applications where a directional antenna isn't required.
2. Directional or *beam* antennas which are intended to preferentially radiate or receive in a particular direction or directional pattern.

In common usage "omnidirectional" usually refers to all horizontal directions, typically with reduced performance in the direction of the sky or the ground (a truly isotropic radiator is not even possible). A "directional" antenna usually is intended to maximize its coupling to the electromagnetic field in the direction of the other station, or sometimes to cover a particular sector such as a 120° horizontal fan pattern in the case of a panel antenna at a cell site.

One example of omnidirectional antennas is the very common *vertical antenna* or whip antenna consisting of a metal rod (often, but not always, a quarter of a wavelength long). A dipole antenna is similar but consists of two such conductors extending in opposite directions, with a total length that is often, but not always, a half of a wavelength long. Dipoles are typically oriented horizontally in which case they are weakly directional: signals are reasonably well radiated toward or received from all directions with the exception of the direction along the conductor itself; this region is called the antenna blind cone or null.

Both the vertical and dipole antennas are simple in construction and relatively inexpensive. The dipole antenna, which is the basis for most antenna designs, is a balanced component, with equal but opposite voltages and currents applied at its two terminals through a balanced transmission line (or to a coaxial transmission line through a so-called balun). The vertical antenna, on the other hand, is a *monopole* antenna. It is typically connected to the inner conductor of a coaxial transmission line (or a matching network); the shield of the transmission line is connected to ground. In this way, the ground (or any large conductive surface) plays the role of the second conductor of a dipole, thereby forming a complete circuit. Since monopole antennas rely on a conductive ground, a so-called grounding structure may be employed in order to provide a better ground contact to the earth or which itself acts as a ground plane to perform that function regardless of (or in absence of) an actual contact with the earth.

Antennas fancier than the dipole or vertical designs are usually intended to increase the directivity and consequently the gain of the antenna. This can be accomplished in many

different ways leading to a plethora of antenna designs. The vast majority of designs are fed with a balanced line (unlike a monopole antenna) and are based on the dipole antenna with additional components (or *elements*) which increase its directionality.

For instance, a phased array consists of two or more simple antennas which are connected together through an electrical network. This often involves a number of parallel dipole antennas with a certain spacing. Depending on the relative phase introduced by the network, the same combination of dipole antennas can operate as a "broadside array" (directional normal to a line connecting the elements) or as an "end-fire array" (directional along the line connecting the elements). Antenna arrays may employ any basic (omnidirectional or weakly directional) antenna type, such as dipole, loop or slot antennas. These elements are often identical.

However a log-periodic dipole array consists of a number of dipole elements of *different* lengths in order to obtain a somewhat directional antenna having an extremely wide bandwidth: these are frequently used for television reception in fringe areas. The dipole antennas composing it are all considered "active elements" since they are all electrically connected together (and to the transmission line). On the other hand, a superficially similar dipole array, the Yagi-Uda Antenna (or simply "Yagi"), has only one dipole element with an electrical connection; the other so-called parasitic elements interact with the electromagnetic field in order to realize a fairly directional antenna but one which is limited to a rather narrow bandwidth. The Yagi antenna has similar looking parasitic dipole elements but which act differently due to their somewhat different lengths. There may be a number of so-called "directors" in front of the active element in the direction of propagation, and usually a single (but possibly more) "reflector" on the opposite side of the active element.

Greater directionality can be obtained using beam-forming techniques such as a parabolic reflector or a horn. Since the size of a directional antenna depends on it being large compared to the wavelength, very directional antennas of this sort are mainly feasible at UHF and microwave frequencies. On the other hand, at low frequencies (such as AM broadcast) where a practical antenna must be much smaller than a wavelength, significant directionality isn't even possible. A vertical antenna or loop antenna small compared to the wavelength is typically used, with the main design challenge being that of impedance matching. With a vertical antenna a *loading coil* at the base of the antenna may be employed to cancel the reactive component of impedance; small loop antennas are tuned with parallel capacitors for this purpose.

An antenna lead-in is the transmission line (or *feed line*) which connects the antenna to a transmitter or receiver. The *antenna feed* may refer to all components connecting the antenna to the transmitter or receiver, such as an impedance matching network in addition to the transmission line. In a so-called aperture antenna, such as a horn or parabolic dish, the "feed" may also refer to a basic antenna inside the entire system (normally at the focus of the parabolic dish or at the throat of a horn) which could be considered the one active element in that antenna system. A microwave antenna may also be fed directly from a waveguide in lieu of a (conductive) transmission line.

An antenna counterpoise or ground plane is a structure of conductive material which improves or substitutes for the ground. It may be connected to or insulated from the natural ground. In a monopole antenna, this aids in the function of the natural ground, particularly where variations (or limitations) of the characteristics of the natural ground interfere with its proper function. Such a structure is normally connected to the return connection of an unbalanced transmission line such as the shield of a coaxial cable.

An electromagnetic wave *refractor* in some aperture antennas is a component which due to its shape and position functions to selectively delay or advance portions of the electromagnetic wavefront passing through it. The refractor alters the spatial characteristics of the wave on one side relative to the other side. It can, for instance, bring the wave to a focus or alter the wave front in other ways, generally in order to maximize the directivity of the antenna system. This is the radio equivalent of an optical lens.

An antenna coupling network is a passive network (generally a combination of inductive and capacitive circuit elements) used for impedance matching in between the antenna and the transmitter or receiver. This may be used to improve the standing wave ratio in order to minimize losses in the transmission line (especially at higher frequencies and/or over longer distances) and to present the transmitter or receiver with a standard resistive impedance (such as 75 ohms) that it expects to see for optimum operation.

Reciprocity

It is a fundamental property of antennas that the characteristics of an antenna described in the next section, such as gain, radiation pattern, impedance, bandwidth, resonant frequency and polarization, are the same whether the antenna is transmitting or receiving. For example, the "*receiving pattern*" (sensitivity as a function of direction) of an antenna when used for reception is identical to the radiation pattern of the antenna when it is *driven* and functions as a radiator. This is a consequence of the reciprocity theorem of electromagnetics. Therefore in discussions of antenna properties no distinction is usually made between receiving and transmitting terminology, and the antenna can be viewed as either transmitting or receiving, whichever is more convenient.

A necessary condition for the above reciprocity property is that the materials in the antenna and transmission medium are linear and reciprocal. *Reciprocal* (or *bilateral*) means that the material has the same response to an electric or magnetic field, or a current, in one direction, as it has to the field or current in the opposite direction. Most materials used in antennas meet these conditions, but some microwave antennas use high-tech components such as isolators and circulators, made of nonreciprocal materials such as ferrite or garnet. These can be used to give the antenna a different behavior on receiving than it has on transmitting, which can be useful in applications like radar.

Parameters

There are several critical parameters affecting an antenna's performance that can be adjusted during the design process. These are resonant frequency, impedance, gain,

aperture or radiation pattern, polarization, efficiency and bandwidth. Transmit antennas may also have a maximum power rating, and receive antennas differ in their noise rejection properties. All of these parameters can be measured through various means.

Resonant frequency

Many types of antenna are tuned to work at one particular frequency, and are effective only over a range of frequencies centered on this frequency, called the resonant frequency. These are called *resonant antennas*. The antenna acts as an electrical resonator. When driven at its resonant frequency, large standing waves of voltage and current are excited in the antenna elements. These large currents and voltages radiate intense electromagnetic waves, so the power radiated by the antenna is maximum at the resonant frequency.

In antennas made of thin linear conductive elements, the length of the driven element(s) determines the resonant frequency. To be resonant, the length of a driven element should typically be either half or a quarter of the wavelength at that frequency; these are called half-wave and quarter-wave antennas. The length referred to is not the physical length, but the electrical length of the element, which is the physical length divided by the velocity factor (the ratio of the speed of wave propagation in the wire to c_0, the speed of light in a vacuum). Antennas are usually also resonant at multiples (harmonics) of the lowest resonant frequency.

Some antenna designs have multiple resonant frequencies, and some are relatively effective over a very broad range of frequencies. or bandwidth. One commonly known type of wide band antenna is the logarithmic or log-periodic antenna.

The resonant frequency also affects the impedance of the antenna. At resonance, the equivalent circuit of an antenna is a pure resistance, with no reactive component. At frequencies other than the resonant frequencies, the antenna has capacitance or inductance as well as resistance. An antenna can be made resonant at other frequencies besides its natural resonant frequency by compensating for these reactances by adding a loading coil or capacitor in series with it. Other properties of an antenna change with frequency, in particular the radiation pattern, so the antenna's operating frequency may be considerably different from the resonant frequency to optimize other important parameters.

Gain

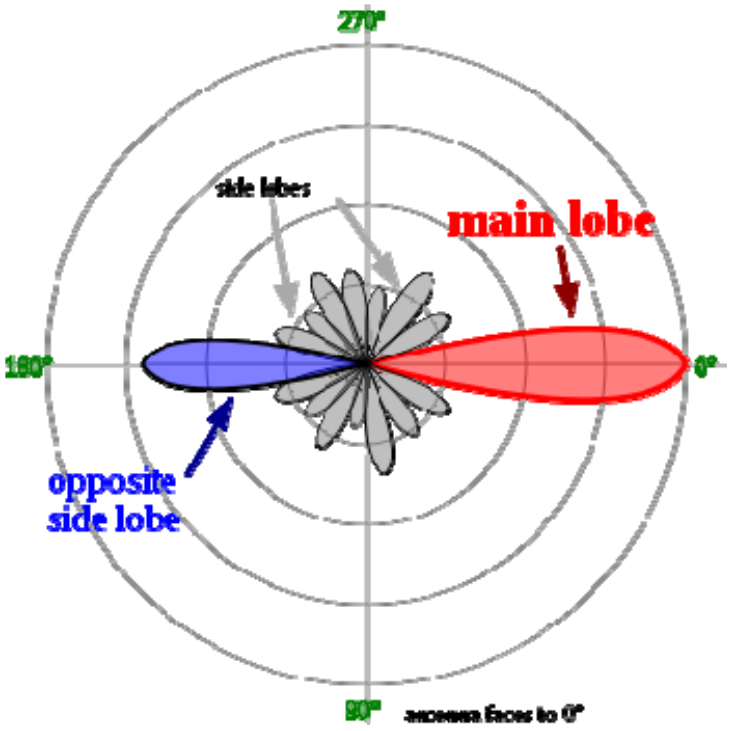
Gain is a parameter which measures the degree of directivity of the antenna's radiation pattern. An antenna with a low gain emits radiation with about the same power in all directions, whereas a high-gain antenna will preferentially radiate in particular directions. Specifically, the *antenna gain*, *directive gain*, or *power gain* of an antenna is defined as the ratio of the intensity (power per unit surface) radiated by the antenna in the direction of its maximum output, at an arbitrary distance, divided by the intensity radiated at the same distance by a hypothetical isotropic antenna.

The gain of an antenna is a passive phenomenon - power is not added by the antenna, but simply redistributed to provide more radiated power in a certain direction than would be transmitted by an isotropic antenna. An antenna designer must take into account the application for the antenna when determining the gain. High-gain antennas have the advantage of longer range and better signal quality, but must be aimed carefully in a particular direction. Low-gain antennas have shorter range, but the orientation of the antenna is relatively inconsequential. For example, a dish antenna on a spacecraft is a high-gain device that must be pointed at the planet to be effective, whereas a typical Wi-Fi antenna in a laptop computer is low-gain, and as long as the base station is within range, the antenna can be in any orientation in space. It makes sense to improve horizontal range at the expense of reception above or below the antenna. Thus most antennas labelled "omnidirectional" really have some gain.

In practice, the half-wave dipole is taken as a reference instead of the isotropic radiator. The gain is then given in **dBd** (decibels over **d**ipole):

NOTE: **0 dBd = 2.15 dBi**. It is vital in expressing gain values that the reference point be included. Failure to do so can lead to confusion and error.

Radiation pattern



polar plots of the horizontal cross sections of a (virtual) Yagi-Uda-antenna. Outline connects points with 3db field power compared to an ISO emitter.

The radiation pattern of an antenna is a plot of the relative field strength of the radio waves emitted by the antenna at different angles. It is typically represented by a three dimensional graph, or polar plots of the horizontal and vertical cross sections. The pattern

of an ideal isotropic antenna, which radiates equally in all directions, would look like a sphere. Many nondirectional antennas, such as monopoles and dipoles, emit equal power in all horizontal directions, with the power dropping off at higher and lower angles; this is called an omnidirectional pattern and when plotted looks like a torus or donut.

The radiation of many antennas shows a pattern of maxima or "*lobes*" at various angles, separated by "*nulls*", angles where the radiation falls to zero. This is because the radio waves emitted by different parts of the antenna typically interfere, causing maxima at angles where the radio waves arrive at distant points in phase, and zero radiation at other angles where the radio waves arrive out of phase. In a directional antenna designed to project radio waves in a particular direction, the lobe in that direction is designed larger than the others and is called the "*main lobe*". The other lobes usually represent unwanted radiation and are called "*sidelobes*". The axis through the main lobe is called the "*principle axis*" or "*boresight axis*".

Impedance

As an electro-magnetic wave travels through the different parts of the antenna system (radio, feed line, antenna, free space) it may encounter differences in impedance (E/H, V/I, etc.). At each interface, depending on the impedance match, some fraction of the wave's energy will reflect back to the source, forming a standing wave in the feed line. The ratio of maximum power to minimum power in the wave can be measured and is called the standing wave ratio (**SWR**). A SWR of 1:1 is ideal. A SWR of 1.5:1 is considered to be marginally acceptable in low power applications where power loss is more critical, although an SWR as high as 6:1 may still be usable with the right equipment. Minimizing impedance differences at each interface (impedance matching) will reduce SWR and maximize power transfer through each part of the antenna system.

Complex impedance of an antenna is related to the electrical length of the antenna at the wavelength in use. The impedance of an antenna can be matched to the feed line and radio by adjusting the impedance of the feed line, using the feed line as an impedance transformer. More commonly, the impedance is adjusted at the load (see below) with an antenna tuner, a balun, a matching transformer, matching networks composed of inductors and capacitors, or matching sections such as the gamma match.

Efficiency

Efficiency is the ratio of power actually radiated to the power put into the antenna terminals. A dummy load may have an SWR of 1:1 but an efficiency of 0, as it absorbs all power and radiates heat but very little RF energy, showing that SWR alone is not an effective measure of an antenna's efficiency. Radiation in an antenna is caused by radiation resistance which can only be measured as part of total resistance including loss resistance. Loss resistance usually results in heat generation rather than radiation, and reduces efficiency. Mathematically, efficiency is calculated as radiation resistance divided by total resistance.

Bandwidth

The *bandwidth* of an antenna is the range of frequencies over which it is effective, usually centered on the resonant frequency. The bandwidth of an antenna may be increased by several techniques, including using thicker wires, replacing wires with *cages* to simulate a thicker wire, tapering antenna components (like in a feed horn), and combining multiple antennas into a single assembly (array) and allowing the natural impedance of suitable inductive RF filter traps to select the correct antenna. All these attempts to increase bandwidth by adding capacitance to the surface area have a detrimental effect on efficiency by reducing the Q factor. They also have an adverse effect on the rejection of unwanted harmonics, on both received and transmitted signal frequencies. Small antennas are usually preferred for convenience, but there is a fundamental limit relating bandwidth, size and efficiency.

Polarization

The *polarization* of an antenna is the orientation of the electric field (E-plane) of the radio wave with respect to the Earth's surface and is determined by the physical structure of the antenna and by its orientation. It has nothing in common with antenna directionality terms: "horizontal", "vertical" and "circular". Thus, a simple straight wire antenna will have one polarization when mounted vertically, and a different polarization when mounted horizontally. "Electromagnetic wave polarization filters" are structures which can be employed to act directly on the electromagnetic wave to filter out wave energy of an undesired polarization and to pass wave energy of a desired polarization.

Reflections generally affect polarization. For radio waves the most important reflector is the ionosphere - signals which reflect from it will have their polarization changed unpredictably. For signals which are reflected by the ionosphere, polarization cannot be relied upon. For line-of-sight communications for which polarization can be relied upon, it can make a large difference in signal quality to have the transmitter and receiver using the same polarization; many tens of dB difference are commonly seen and this is more than enough to make the difference between reasonable communication and a broken link.

Polarization is largely predictable from antenna construction but, especially in directional antennas, the polarization of side lobes can be quite different from that of the main propagation lobe. For radio antennas, polarization corresponds to the orientation of the radiating element in an antenna. A vertical omnidirectional WiFi antenna will have vertical polarization (the most common type). An exception is a class of elongated waveguide antennas in which vertically placed antennas are horizontally polarized. Many commercial antennas are marked as to the polarization of their emitted signals.

Polarization is the sum of the E-plane orientations over time projected onto an imaginary plane perpendicular to the direction of motion of the radio wave. In the most general case, polarization is elliptical, meaning that the polarization of the radio waves varies over time. Two special cases are linear polarization (the ellipse collapses into a line) and

circular polarization (in which the two axes of the ellipse are equal). In linear polarization the antenna compels the electric field of the emitted radio wave to a particular orientation. Depending on the orientation of the antenna mounting, the usual linear cases are horizontal and vertical polarization. In circular polarization, the antenna continuously varies the electric field of the radio wave through all possible values of its orientation with regard to the Earth's surface. Circular polarizations, like elliptical ones, are classified as right-hand polarized or left-hand polarized using a "thumb in the direction of the propagation" rule. Optical researchers use the same rule of thumb, but pointing it in the direction of the emitter, not in the direction of propagation, and so are opposite to radio engineers' use.

In practice, regardless of confusing terminology, it is important that linearly polarized antennas be matched, lest the received signal strength be greatly reduced. So horizontal should be used with horizontal and vertical with vertical. Intermediate matchings will lose some signal strength, but not as much as a complete mismatch. Transmitters mounted on vehicles with large motional freedom commonly use circularly polarized antennas so that there will never be a complete mismatch with signals from other sources.

Transmission and reception

All of the antenna parameters are expressed in terms of a transmission antenna, but are identically applicable to a receiving antenna, due to reciprocity. Impedance, however, is not applied in an obvious way; for impedance, the impedance at the load (where the power is consumed) is most critical. For a transmitting antenna, this is the antenna itself. For a receiving antenna, this is at the (radio) receiver rather than at the antenna. Tuning is done by adjusting the length of an electrically long linear antenna to alter the electrical resonance of the antenna.

Antenna tuning is done by adjusting an inductance or capacitance combined with the active antenna (but distinct and separate from the active antenna). The inductance or capacitance provides the reactance which combines with the inherent reactance of the active antenna to establish a resonance in a circuit including the active antenna. The established resonance being at a frequency other than the natural electrical resonant frequency of the active antenna. Adjustment of the inductance or capacitance changes this resonance.

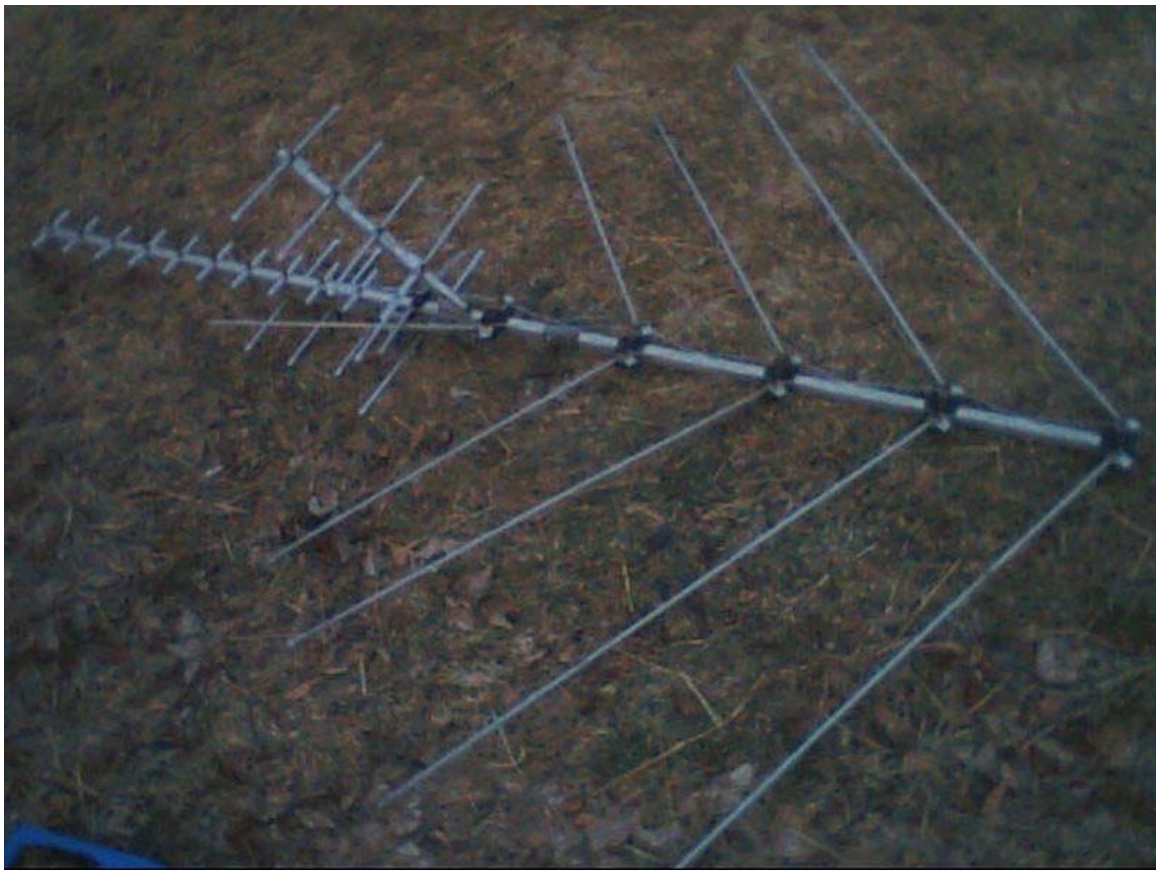
Antennas used for transmission have a maximum power rating, beyond which heating, arcing or sparking may occur in the components, which may cause them to be damaged or destroyed. Raising this maximum power rating usually requires larger and heavier components, which may require larger and heavier supporting structures. This is a concern only for transmitting antennas, as the power received by an antenna rarely exceeds the microwatt range.

Antennas designed specifically for reception might be optimized for noise rejection capabilities. An *antenna shield* is a conductive or low reluctance structure (such as a wire, plate or grid) which is adapted to be placed in the vicinity of an antenna to reduce,

as by dissipation through a resistance or by conduction to ground, undesired electromagnetic radiation, or electric or magnetic fields, which are directed toward the active antenna from an external source or which emanate from the active antenna. Other methods to optimize for noise rejection can be done by selecting a narrow bandwidth so that noise from other frequencies is rejected, or selecting a specific radiation pattern to reject noise from a specific direction, or by selecting a polarization different from the noise polarization, or by selecting an antenna that favors either the electric or magnetic field.

For instance, an antenna to be used for reception of low frequencies (below about ten megahertz) will be subject to both man-made noise from motors and other machinery, and from natural sources such as lightning. Successfully rejecting these forms of noise is an important antenna feature. A small coil of wire with many turns is more able to reject such noise than a vertical antenna. However, the vertical will radiate much more effectively on transmit, where extraneous signals are not a concern.

Basic antenna models



Typical US multiband TV antenna (aerial)

There are many variations of antennas. Below are a few basic models. More can be found in Category:Radio frequency antenna types.

- The isotropic radiator is a purely theoretical antenna that radiates equally in all directions. It is considered to be a point in space with no dimensions and no mass. This antenna cannot physically exist, but is useful as a theoretical model for comparison with all other antennas. Most antennas' gains are measured with reference to an isotropic radiator, and are rated in dBi (decibels with respect to an isotropic radiator).
- The dipole antenna is simply two wires pointed in opposite directions arranged either horizontally or vertically, with one end of each wire connected to the radio and the other end hanging free in space. Since this is the simplest practical antenna, it is also used as a reference model for other antennas; gain with respect to a dipole is labeled as dBd. Generally, the dipole is considered to be omnidirectional in the plane perpendicular to the axis of the antenna, but it has deep nulls in the directions of the axis. Variations of the dipole include the folded dipole, the half wave antenna, the ground plane antenna, the whip, and the J-pole.
- The Yagi-Uda antenna is a directional variation of the dipole with parasitic elements added which are functionality similar to adding a reflector and lenses (directors) to focus a filament light bulb.
- The random wire antenna is simply a very long (at least one quarter wavelength) wire with one end connected to the radio and the other in free space, arranged in any way most convenient for the space available. Folding will reduce effectiveness and make theoretical analysis extremely difficult. (The added length helps more than the folding typically hurts.) Typically, a random wire antenna will also require an antenna tuner, as it might have a random impedance that varies non-linearly with frequency.
- The horn is used where high gain is needed, the wavelength is short (microwave) and space is not an issue. Horns can be narrow band or wide band, depending on their shape. A horn can be built for any frequency, but horns for lower frequencies are typically impractical. Horns are also frequently used as reference antennas.
- The parabolic antenna consists of an active element at the focus of a parabolic reflector to reflect the waves into a plane wave. Like the horn it is used for high gain, microwave applications, such as satellite dishes.
- The patch antenna consists mainly of a square conductor mounted over a groundplane. Another example of a planar antenna is the tapered slot antenna (TSA), as the Vivaldi-antenna.

Practical antennas

"Rabbit ears" set-top antenna

Although any circuit can radiate if driven with a signal of high enough frequency, most practical antennas are specially designed to radiate efficiently at a particular frequency. An example of an inefficient antenna is the simple Hertzian dipole antenna, which radiates over wide range of frequencies and is useful for its small size. A more efficient variation of this is the half-wave dipole, which radiates with high efficiency when the signal wavelength is twice the electrical length of the antenna.

One of the goals of antenna design is to minimize the reactance of the device so that it appears as a resistive load. An "antenna inherent reactance" includes not only the distributed reactance of the active antenna but also the natural reactance due to its location and surroundings (as for example, the capacity relation inherent in the position of the active antenna relative to ground). Reactance diverts energy into the reactive field, which causes unwanted currents that heat the antenna and associated wiring, thereby wasting energy without contributing to the radiated output. Reactance can be eliminated by operating the antenna at its resonant frequency, when its capacitive and inductive reactances are equal and opposite, resulting in a net zero reactive current. If this is not possible, compensating inductors or capacitors can instead be added to the antenna to cancel its reactance as far as the source is concerned.

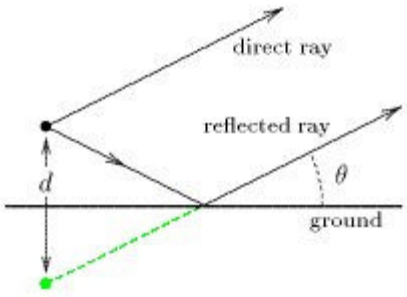
Once the reactance has been eliminated, what remains is a pure resistance, which is the sum of two parts: the ohmic resistance of the conductors, and the radiation resistance. Power absorbed by the ohmic resistance becomes waste heat, and that absorbed by the radiation resistance becomes radiated electromagnetic energy. The greater the ratio of radiation resistance to ohmic resistance, the more efficient the antenna.

Effect of ground

Antennas are typically used in an environment where other objects are present that may have an effect on their performance. Height above ground has a very significant effect on the radiation pattern of some antenna types.

At frequencies used in antennas, the ground behaves mainly as a dielectric. The conductivity of ground at these frequencies is negligible. When an electromagnetic wave arrives at the surface of an object, two waves are created: one enters the dielectric and the other is reflected. If the object is a conductor, the transmitted wave is negligible and the reflected wave has almost the same amplitude as the incident one. When the object is a dielectric, the fraction reflected depends (among others things) on the angle of incidence. When the angle of incidence is small (that is, the wave arrives almost perpendicularly) most of the energy traverses the surface and very little is reflected. When the angle of incidence is near 90° (grazing incidence) almost all the wave is reflected.

Most of the electromagnetic waves emitted by an antenna to the ground below the antenna at moderate (say < 60°) angles of incidence enter the earth and are absorbed (lost). But waves emitted to the ground at grazing angles, far from the antenna, are almost totally reflected. At grazing angles, the ground behaves as a mirror. Quality of reflection depends on the nature of the surface. When the irregularities of the surface are smaller than the wavelength reflection is good.



The wave reflected by earth can be considered as emitted by the image antenna

This means that the receptor "sees" the real antenna and, under the ground, the image of the antenna reflected by the ground. If the ground has irregularities, the image will appear fuzzy.

If the receiver is placed at some height above the ground, waves reflected by ground will travel a little longer distance to arrive to the receiver than direct waves. The distance will be the same only if the receiver is close to ground.

In the drawing at right, we have drawn the angle θfar bigger than in reality. Distance between the antenna and its image is d.

The situation is a bit more complex because the reflection of electromagnetic waves depends on the polarization of the incident wave. As the refractive index of the ground (average value $\simeq 2$) is bigger than the refractive index of the air ($\simeq 1$), the direction of the component of the electric field parallel to the ground inverses at the reflection. This is equivalent to a phase shift of πradians or 180°. The vertical component of the electric field reflects without changing direction. This sign inversion of the parallel component and the non-inversion of the perpendicular component would also happen if the ground were a good electrical conductor.

The vertical component of the current reflects without changing sign. The horizontal component reverses sign at reflection.

This means that a receiving antenna "sees" the image antenna with the current in the same direction if the antenna is vertical or with the current inverted if the antenna is horizontal.

For a vertical polarized emission antenna the far electric field of the electromagnetic wave produced by the direct ray plus the reflected ray is:

$$|E_{\perp}| = 2\,|E_{\theta_1}|\,\left|\cos\left(\tfrac{kd}{2}\sin\theta\right)\right|$$

The sign inversion for the parallel field case just changes a cosine to a sine:

$$|E_{\parallel}| = 2\,|E_{\theta_1}|\,\left|\sin\left(\tfrac{kd}{2}\sin\theta\right)\right|$$

In these two equations:

- $E_{\theta 1}$is the electrical field radiated by the antenna if there were no ground.
- $k=\frac{2\pi}{\lambda}$is the wave number.
- $\lambda$is the wave length.

- d is the distance between antenna and its image (twice the height of the center of the antenna).

Radiation patterns of antennas and their images reflected by the ground. At left the polarization is vertical and there is always a maximum for $\theta=0$. If the polarization is horizontal as at right, there is always a zero for $\theta=0$.

For emitting and receiving antenna situated near the ground (in a building or on a mast) far from each other, distances traveled by direct and reflected rays are nearly the same. There is no induced phase shift. If the emission is polarized vertically the two fields (direct and reflected) add and there is maximum of received signal. If the emission is polarized horizontally the two signals subtracts and the received signal is minimum. This is depicted in the image at right. In the case of vertical polarization, there is always a maximum at earth level (left pattern). For horizontal polarization, there is always a minimum at earth level. Note that in these drawings the ground is considered as a perfect mirror, even for low angles of incidence. In these drawings the distance between the antenna and its image is just a few wavelengths. For greater distances, the number of lobes increases.

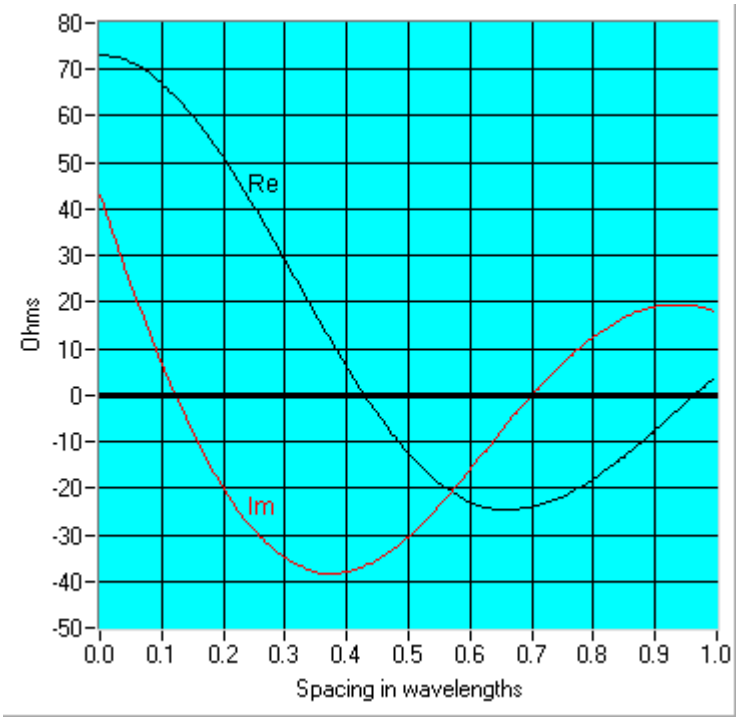
Note that the situation is different–and more complex–if reflections in the ionosphere occur. This happens over very long distances (thousands of kilometers). There is not a direct ray but several reflected rays that add with different phase shifts.

This is the reason why almost all public address radio emissions have vertical polarization. As public users are near ground, horizontal polarized emissions would be poorly received. Observe household and automobile radio receivers. They all have vertical antennas or horizontal ferrite antennas for vertical polarized emissions. In cases where the receiving antenna must work in any position, as in mobile phones, the emitter and receivers in base stations use circular polarized electromagnetic waves.

Classical (analog) television emissions are an exception. They are almost always horizontally polarized, because the presence of buildings makes it unlikely that a good emitter antenna image will appear. However, these same buildings reflect the electromagnetic waves and can create ghost images. Using horizontal polarization, reflections are attenuated because of the low reflection of electromagnetic waves whose magnetic field is parallel to the dielectric surface near the Brewster's angle. Vertically polarized analog television has been used in some rural areas. In digital terrestrial

television reflections are less obtrusive, due to the inherent robustness of digital signalling and built-in error correction.

Mutual impedance and interaction between antennas



Mutual impedance between parallel $\frac{\lambda}{2}$dipoles not staggered. Curves **Re** and **Im** are the resistive and reactive parts of the impedance.

Current circulating in any antenna induces currents in all others. One can postulate a **mutual impedance** Z_{12}between two antennas that has the same significance as the $j\omega M$in ordinary coupled inductors. The mutual impedance Z_{12}between two antennas is defined as:

$$Z_{12} = \frac{v_2}{i_1}$$

where $i_1$is the current flowing in antenna 1 and $v_2$is the voltage that would have to be applied to antenna 2–with antenna 1 removed–to produce the current in the antenna 2 that was produced by antenna 1.

From this definition, the currents and voltages applied in a set of coupled antennas are:

$$
\begin{aligned}
v_1 &= i_1 Z_{11} + i_2 Z_{12} + \cdots + i_n Z_{1n} \\
v_2 &= i_1 Z_{21} + i_2 Z_{22} + \cdots + i_n Z_{2n} \\
\vdots &\quad\ \vdots \quad\quad\ \vdots \quad\quad\quad\quad\ \vdots \\
v_n &= i_1 Z_{n1} + i_2 Z_{n2} + \cdots + i_n Z_{nn}
\end{aligned}
$$

where:

- v_i is the voltage applied to the antenna *i*
- Z_{ii} is the impedance of antenna *i*
- Z_{ij} is the mutual impedance between antennas *i* and *j*

Note that, as is the case for mutual inductances,

$$Z_{ij} = Z_{ji}$$

This is a consequence of Lorentz reciprocity. If some of the elements are not fed (there is a short circuit instead a feeder cable), as is the case in television antennas (Yagi-Uda antennas), the corresponding v_i are zero. Those elements are called parasitic elements. Parasitic elements are unpowered elements that either reflect or absorb and reradiate RF energy.

In some geometrical settings, the mutual impedance between antennas can be zero. This is the case for crossed dipoles used in circular polarization antennas.

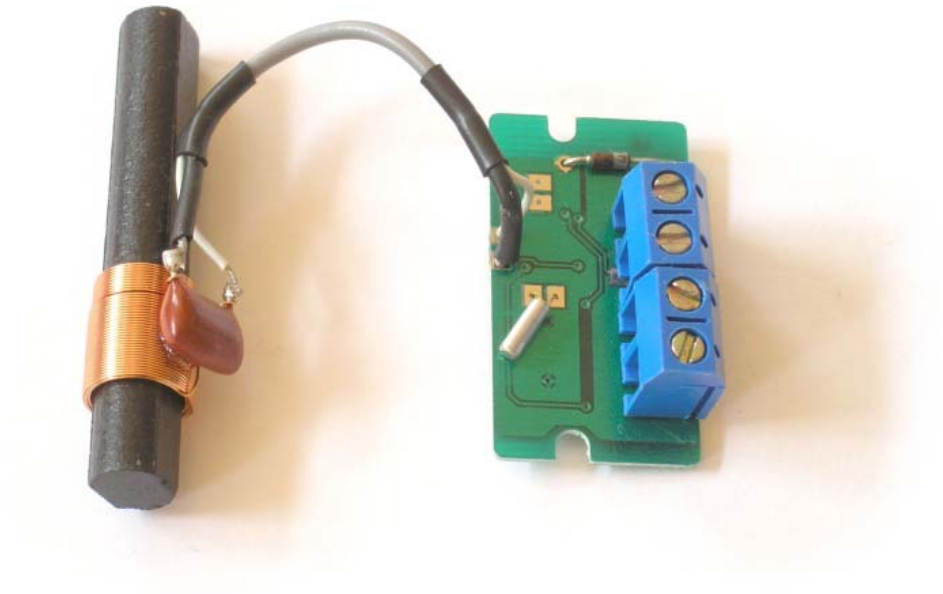
Antenna

Antennas and antenna arrays

A Yagi-Uda beam antenna.

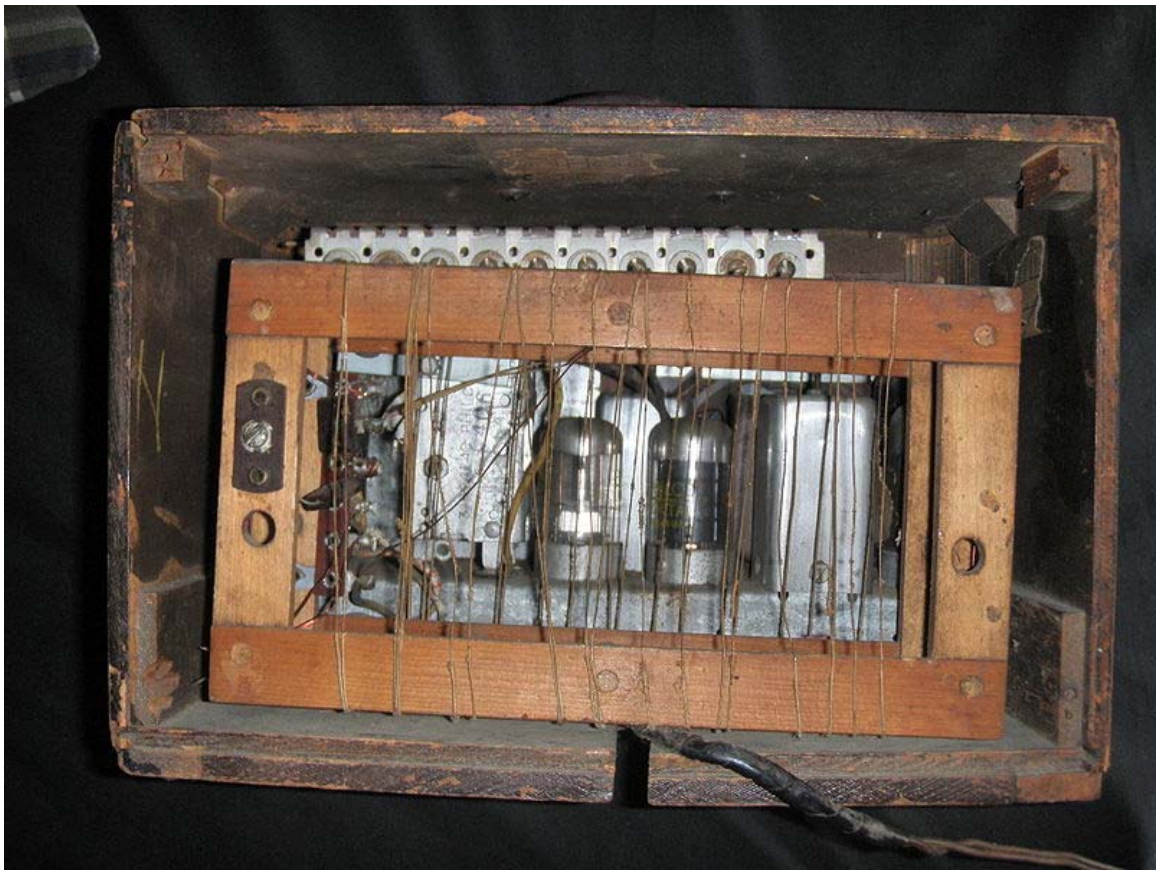
Rooftop TV antenna. It is actually three Yagi antennas. The longest elements are for the low band, while the medium and short elements are for the high and UHF band.

Examples of US 136-174 MHz base station antennas.



Low cost LF time signal receiver, antenna (left) and receiver (right).

"Rabbit ears" antenna



AM loop antenna

Antennas and supporting structures

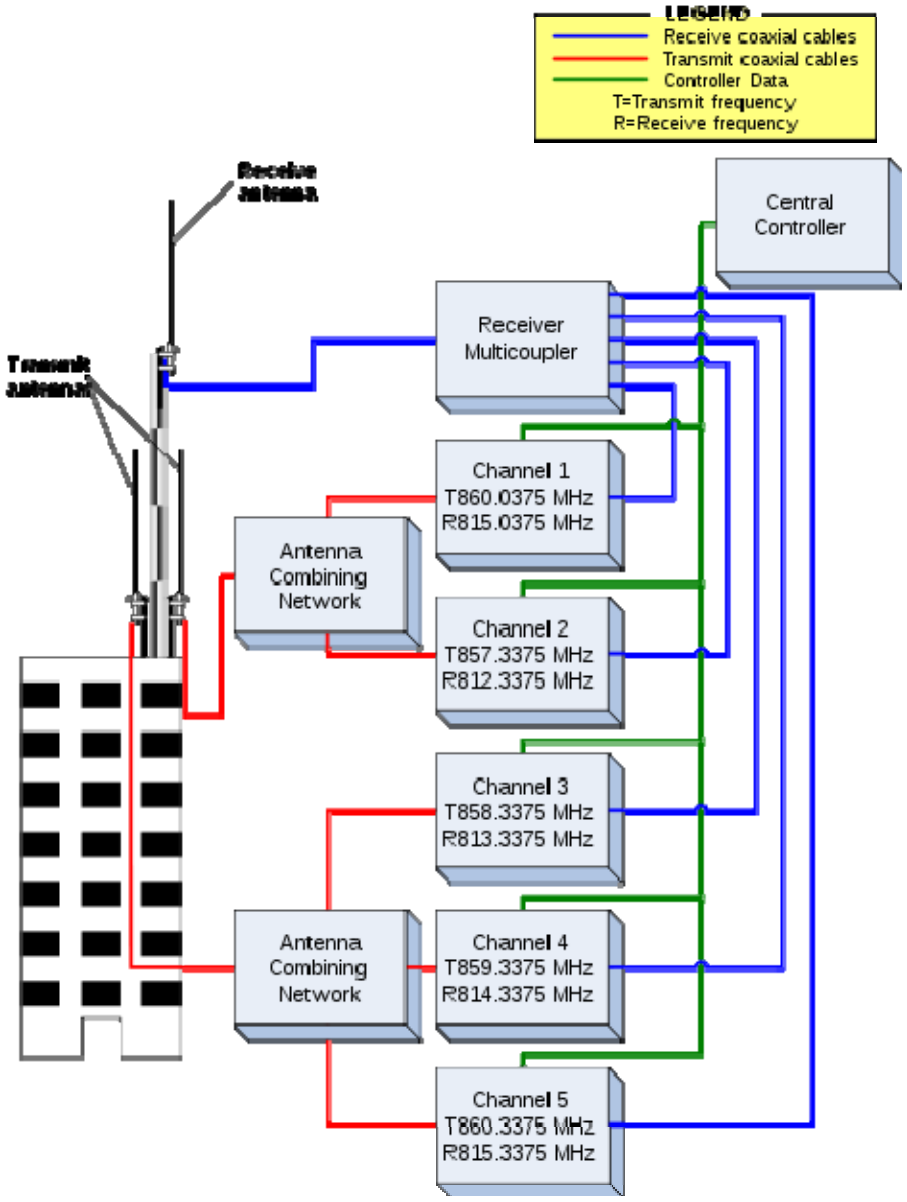
A building rooftop supporting numerous dish and sectored mobile telecommunications antennas (Doncaster, Victoria, Australia).

A water tower in Palmerston, Northern Territory with radio broadcasting and communications antennas.

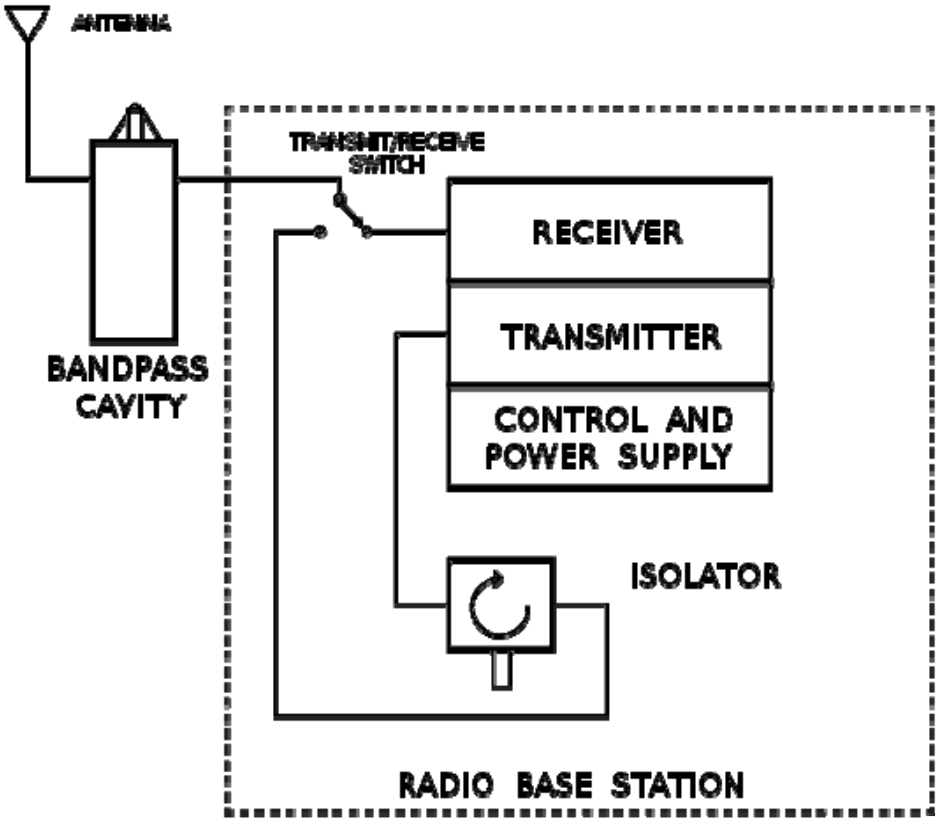
A three-sector telephone site in Mexico City.

Telephone site concealed as a palm tree.

Diagrams as part of a system



Antennas may be connected through a multiplexing arrangement in some applications like this trunked two-way radio example.



Antenna network for an emergency medical services base station.

Chapter 6

Television Antenna

A Winegard HD-7084P 68 element VHF/UHF aerial antenna

A **television antenna**, or **TV aerial**, is an antenna specifically designed for the reception of over the air broadcast television signals, which are transmitted at frequencies from about 41 to 250 MHz in the VHF band, and 470 to 960 MHz in the UHF band in different countries. To cover this range antennas generally consist of multiple conductors of

different lengths which correspond to the wavelength range the antenna is intended to receive. The length of the elements of a TV antenna are usually half the wavelength of the signal they are intended to receive. The wavelength of a signal equals the speed of light (c) divided by the frequency. The design of a television broadcast receiving antenna is the same for the older analog transmissions and the digital television transmissions which are replacing them Sellers often claim to supply a special "digital" or "high definition" antenna advised as a replacement for an existing analog antenna, even if satisfactory: this is misinformation to generate sales of unneeded equipment.

Simple/indoor

Very common "rabbit ears" set-top antenna of older model

Simple half-wave dipole VHF antennas or UHF loop antennas that are made to be placed indoors are often used for television (and VHF radio); these are often called "rabbit ears" or "bunny aerials". because of their appearance. The length of the telescopic "ears" can be adjusted by the user, and should be about one half of the wavelength of the signal for the desired channel. These are not as efficient as an aerial rooftop antenna since they are less directional and not always adjusted to the proper length for the desired channel. Dipole antennas are bi-directional, that is, they receive evenly forward and backwards, and also cover a broader band than antennas with more elements. This makes them less efficient than antennas designed to maximise the signal from a narrower angle in one direction.

Coupled with the poor placing, indoors and closer to the ground, they are much worse than multi-element rooftop antennas at receiving signals which are not very strong, although often adequate for nearby transmitters, in which case they may be adequate and cheap. These simple antennas are called set-top antennas because they are often placed on top of the television set or receiver.

The actual length of the ears is optimally about 91% of half the wavelength of the desired channel in free space. Quarter-wave television antennas are also used. These use a single element, and use the earth as a ground plane; therefore, no ground is required in the feed line.*a*

Outdoor

An aerial or rooftop antenna generally consists of multiple conductive elements that are arranged such that it is a directional antenna. The length of the elements is about one half of the signal wavelength. Therefore, the length of each element corresponds to a certain frequency.

In a combined VHF/UHF antenna the longer elements (for picking up VHF frequencies) are at the "back" of the antenna, relative to the device's directionality, and the much shorter UHF elements are in the "front", and the antenna works best when "pointing" to

the source of the signal to be received. The smallest elements in this design, located in the "front", are UHF director elements, which are usually identical and give the antenna its directionality, as well as improving gain. The longest elements, located in the "back" of the antenna form a VHF phased array. Other long elements may be UHF reflectors Another common aerial antenna element is the corner reflector, a type of UHF reflector which increases gain and directionality for UHF frequencies.

An antenna can have a smaller or larger number of directors; the more directors it has (requiring a longer boom), and the more accurate their tuning, the higher its gain will be. For the commonly used Yagi antenna this is not a linear relationship. Antenna gain is the ratio of the signal received from the preferred direction to the signal from an ideal omnidirectional antenna. Gain is inversely proportional to the antenna's acceptance angle. The thickness of the rods on a Yagi antenna and its bandwidth are inversely proportional; thicker rods provide a wider band. Thinner rods are preferable to provide a narrower band, hence higher gain in the preferred direction; however, they must be thick enough to withstand wind.

Two or more directional rooftop antennas can be set up and connected to one receiver. Antennas designed for rooftop use are sometimes located in attics.

Sometimes television transmitters are organised such that all receivers in a given location need receive transmissions in only a relatively narrow band of the full UHF television spectrum and from the same direction, so that a single antenna provides reception from all stations.

Types of outdoor antenna

A UHF television antenna

An antenna pole setup in a chimney, reaching 35 feet (10.7 meters) off the ground

Small multi-directional: The smallest of all outdoor television antennas. They are designed to receive equal amounts of signal from all directions. These generally receive signals up to a maximum of thirty miles away from the transmitting station, greatly depending on the type. But, things such as large buildings or thick woods may greatly affect signal. They come in many different styles, ranging from small dishes to small metal bars, some can even mount on existing satellite dishes.

Medium multi-directional: A step up from the small multi-directional, these also receive signals from all directions. These usually require an amplifier in situations when long cable lengths are between the television receiver and the antenna. Styles are generally similar to small multi-directionals, but slightly larger.

Large multi-directional: These are the largest of all multi-directional outdoor television antennas. Styles include large "nets" or dishes, but can also greatly vary. Depending on the type, signal reception usually ranges from 30 to up to 70 miles.

Small directional: The smallest of all directional antennas, these antennas are multi-element antennas, typically placed on rooftops. This style of antenna receives signals generally equal to that of large multi-directionals. One advantage that small directionals hold, however, is that they can significantly reduce "ghosting" effects of television picture.

Medium directional: These antennas are the ones most often seen on suburban rooftops. Usually consisting of many elements, and slightly larger than the small directionals, these antennas are ideal for receiving television signals in suburban areas. Signal usually ranges from 30 to 60 miles away from the broadcasting station.

Large directional: The largest of all common outdoor television antennas, these antennas are designed to receive the weakest available stations in an area. Larger than the medium directional, this type of antenna consists of many elements and is usually used in rural areas, where reception is difficult. When used in conjunction with an amplifier, these antennas can usually pick up stations from 60 up to and over 100 miles, depending on the type.

The use of outdoor antennas with an amplifier can improve signal on low signal strength channels. If the signal quality is low repositioning the antenna onto a high mast will improve signal

Installation

A short antenna pole next to a house; this setup would only work well for receiving signals on that side of the house as they would not go through stone, especially.

Multiple Yagi TV aerials in Israel

Antennas are commonly placed on rooftops, and sometimes in attics. Placing an antenna indoors significantly attenuates the signal available to it. Directional antennas must be pointed at the transmitter they are receiving; in most cases great accuracy is not needed. In a given region it is sometimes arranged that all television transmitters are located in roughly the same direction and use frequencies space closely enough that a single antenna suffices for all. A single transmitter location may transmit signals for several channels.

Analog television signals are susceptible to ghosting in the image, multiple closely-spaced images giving the impression of blurred and repeated images of edges in the picture. This was due to the signal being reflected from nearby objects (buildings, tree,

mountains); several copies of the signal, of different strengths and subject to different delays, are picked up. This was different for different transmissions. Careful positioning of the antenna could produce a compromise position which minimized the ghosts on different channels. Ghosting is also possible if multiple antennas connected to the same receiver pick up the same station, especially if the lengths of the cables connecting them to the splitter/merger were different lengths or the antennas were too close together. Analog television is being replaced by digital, which is not subject to ghosting.

Rooftop and other outdoor antennas

Aerials are attached to roofs in various ways, usually on a pole to elevate it above the roof. This is generally sufficient in most areas. In some places; however, such as a deep valley or near taller structures, the antenna may need to be placed significantly higher, using a lattice tower or mast.

The higher the antenna is placed, the better it will perform. An antenna of higher gain will be able to receive weaker signals from its preferred direction. Intervening buildings, topographical features (mountains), and dense forest will weaken the signal; in many cases the signal will be reflected such that a usable signal is still available. There are physical dangers inherent to high or complex antennas, such as the structure falling or being destroyed by the weather. There are also varying local ordinances which restrict and limit such things as the height of a structure without obtaining permits. For example, in the USA, the Telecommunications Act of 1996 allows any homeowner to install "An antenna that is designed to receive local television broadcast signals", but that "masts higher than 12 feet above the roof-line may be subject to local permitting requirements."

Indoor antennas

As discussed previously, antennas may be placed indoors where signals are strong enough to overcome antenna shortcomings. The antenna is simply plugged into the television receiver and placed conveniently, often on the top of the receiver ("set-top"). Sometimes the position needs to be experimented with to get the best picture. Indoor antennas can also benefit from RF amplification, commonly called a TV booster. Indoor antennas will never be an option in weak signal areas.

Attic installation

Sometimes it is desired not to put an antenna on the roof; in these cases, antennas designed for outdoor use are often mounted in the attic or loft, although antennas designed for attic use are also available. Putting an antenna indoors significantly decreases its performance due to lower elevation above ground level and intervening walls; however, in strong signal areas reception may be satisfactory. One layer of asphalt shingles, roof felt, and a plywood roof deck are considered to attenuate the signal to about half.

Multiple antennas, rotators

Two aerials setup on a roof. Spaced horizontally and vertically

It is sometimes desired to receive signals from transmitters which are not in the same direction. This can be achieved, for one station at a time, by using a rotator operated by an electric motor to turn the antenna as desired. Alternatively, two or more antennas, each pointing at a desired transmitter and coupled by appropriate circuitry, can be used. To prevent the antennas interfering with each other, the vertical spacing between the booms must be at least half the wavelength of the lowest frequency to be received (Distance=λ/2). The wavelength of 54 MHz (Channel 2) is 5.5 meters ($\lambda \times f = c$) so the antennas must be a minimum of 2.25 meters, or ~89 inches apart. It is also important that the cables connecting the antennas to the signal splitter/merger be exactly the same length, to prevent phasing issues,which cause ghosting with analog reception. That is, the antennas might both pick up the same station; the signal from the one with the shorter cable will reach the receiver slightly sooner, supplying the receiver with two pictures slightly offset. There may be phasing issues even with the same length of down-lead cable. Bandpass filters or "signal traps" may help to reduce this problem.

For side-by-side placement of multiple antennas, as is common in a space of limited height such as an attic, they should be separated by at least one full wavelength of the lowest frequency to be received at their closest point.

Often when multiple antennas are used, one is for a range of co-located stations and the other is for a single transmitter in a different direction.

Safety

- TV antennas are large conductors of electricity and attract lightning, acting as a lightning rod. The use of a lightning arrestor is usual to protect against this. A large grounding rod connected to both the antenna and the mast or pole is required.

- Properly installed masts, especially tall ones, are guyed with galvanized cable; no insulators are needed. They are designed to withstand worst-case weather conditions in the area, and positioned so that they do not interfere with power lines if they fall.

- There is inherent danger in being on the rooftop of a house, required for installing or adjusting a television antenna.

Chapter 7

Radio Masts and Towers

Masts of the Rugby VLF transmitter in England

A dismantled radio mast in sections

Radio masts and towers are, typically, tall structures designed to support antennas (also known as aerials) for telecommunications and broadcasting, including television. They are among the tallest man-made structures. Similar structures include electricity pylons and towers for wind turbines.

Masts are sometimes named after the broadcasting organisations that use them, or after a nearby city or town.

The Warsaw Radio Mast was the world's tallest supported structure on land, but it collapsed in 1991, leaving the KVLY/KTHI-TV mast as the tallest.

In the case of a mast radiator or radiating tower, the whole mast or tower is itself the transmitting antenna.

Mast or tower?

A radio mast base showing how virtually all support is provided by the guy-wires

The terms "mast" and "tower" are often used interchangeably. However, in structural engineering terms, a tower is a self-supporting or cantilevered structure, while a mast is held up by stays or guys. By contrast, in broadcast engineering, a tower is an antenna structure attached to the ground, whereas a mast is a vertical antenna support mounted on some other structure (which itself may be a tower, a building, or a vehicle). Masts (to use the civil engineering terminology) tend to be cheaper to build but require an extended area surrounding them to accommodate the guy wires. Towers are more commonly used in cities where land is in short supply.

There are a few borderline designs which are partly free-standing and partly guyed. For example:

- The Gerbrandy tower consists of a self-supporting tower with a guyed mast on top.
- The few remaining Blaw-Knox towers do the opposite: they have a guyed lower section surmounted by a freestanding part.
- Zendstation Smilde a tall tower with a guyed mast on top (guys go to ground)

- Torre de Collserola a guyed tower, with a guyed mast on top. (Tower portion is not free-standing.)

Materials

Steel lattice

Steel lattice tower

The steel lattice is the most widespread form of construction. It provides great strength, low weight and wind resistance, and economy in the use of materials. Lattices of triangular cross-section are most common, and square lattices are also widely used.

When built as a stayed mast, usually the whole mast is parallel-sided. One exception is the Blaw-Knox type.

When built as a tower, the structure may be parallel-sided or taper over part or all of its height. When constructed of several sections which taper exponentially with height, in the manner of the Eiffel Tower, the tower is said to be an Eiffelized one. The Crystal Palace tower in London is an example.

Tubular steel

Guyed masts are sometimes also constructed out of steel tubes. This construction type has the advantage that cables and other equipment is protected from weather influence and that the structure may look nicer. They are mainly used for FM-/TV-broadcasting, but sometimes also as mast radiator, wherefore the big mast of Mühlacker transmitting station is a good example. A disadvantage of this mast type is that it is much more affected by winds than masts with open bodies. In fact several tubular guyed masts collapsed: in the UK, these were masts the Emley Moor and Waltham TV stations, which collapsed in the 1960s, in Germany that of Bielstein transmitter, which collapsed in 1985. Not in all countries such masts were built: while in Germany, France, UK, Czech, Slovakia and the former Soviet Union multiple tubular guyed masts were built, there are nearly none in Poland and North America.

At several cities in Russia and Ukraine, between 1960 and 1965 several tubular guyed masts with crossbars running from the mast structure to the guys were built. All these masts are exclusively used for FM and TV transmission and are except of the mast in Vinnytsia between 150 and 200 metres tall. The crossbars of these masts are equipped with a gangway and are equipped with smaller antennas. Their main purpose is oscillation damping.

First modern TV Tower in Stuttgart

Reinforced concrete

Reinforced concrete towers are relatively expensive to build but provide a high degree of mechanical rigidity in strong winds. This can be important when antennas with narrow beamwidths are used, such as those used for microwave point-to-point links, and when the structure is to be occupied by people.

In the 1950s, AT&T built numerous concrete towers, more resembling silos than towers, for its first transcontinental microwave route. Many are still in use today.

In Germany and the Netherlands most towers constructed for point-to-point microwave links are built of reinforced concrete, while in the UK most are lattice towers.

Concrete towers can form prestigious landmarks, such as the CN Tower in Toronto. As well as accommodating technical staff, these buildings may have public areas such as observation decks or restaurants.

The Stuttgart TV tower was the first tower in the world to be built in reinforced concrete. It was designed in 1956 by the local civil engineer Fritz Leonhardt.

Tokyo Tower

Fibreglass

Fibreglass poles are occasionally used for low-power non-directional beacons or medium-wave broadcast transmitters.

Wood

There are fewer wooden towers now than in the past. Many were built in the UK during World War II because of a shortage of steel. In Germany before World War II wooden towers were used at nearly all medium-wave transmission sites, but all of these towers have since been demolished, except for the Gliwice Radio Tower.

Ferryside Relay is an example of a TV relay transmitter using a wooden pole.

Other types of antenna supports and structures

Poles

Shorter masts may consist of a self-supporting or guyed wooden pole, similar to a telegraph pole. Sometimes self-supporting tubular galvanized steel poles are used: these may be termed monopoles.

Buildings

In some cases, it is possible to install transmitting antennas on the roofs of tall buildings. In North America, for instance, there are transmitting antennas on the Empire State Building, the Willis Tower ,and formerly on the World Trade Center towers. When the buildings collapsed, several local TV and radio stations were knocked off the air until backup transmitters could be put into service. Such facilities also exist in Europe, particularly for portable radio services and low-power FM radio stations.

Disguised cell-sites

Completed in December 2009 at Epiphany Lutheran Church in Lake Worth, Florida, this 100' tall cross conceals equipment for T-Mobile.

Many people view bare cellphone towers as ugly and an intrusion into their neighbourhoods. Even though people increasingly depend upon cellular communications, they are opposed to the bare towers spoiling otherwise scenic views. Many companies offer to 'hide' cellphone towers in, or as, trees, church towers, flag poles, water tanks and other features. There are many providers that offer these services as part of the normal tower installation and maintenance service. These are generally called "stealth towers" or "stealth installations".

The level of detail and realism achieved by disguised cellphone towers is remarkably high; for example, such towers disguised as trees are nearly indistinguishable from the real thing, even for local wildlife (who additionally benefit from the artificial flora). Such towers can be placed unobtrusively in national parks and other such protected places, such as towers disguised as cacti in Coronado National Forest.

Even when disguised, however, such towers can create controversy; a tower doubling as a flagpole attracted controversy in 2004 in relation to the U.S. Presidential campaign of that year, and highlighted the sentiment that such disguises serve more to allow the installation of such towers in subterfuge away from public scrutiny rather than to serve towards the beautification of the landscape.

Mast radiators

A mast radiator is a radio tower or mast in which the whole structure works as an antenna. It is used frequently as a transmitting antenna for long or medium wave broadcasting.

Structurally, the only difference is that a mast radiator may be supported on an insulator at its base. In the case of a tower, there will be one insulator supporting each leg.

Telescopic, pump-up and tiltover towers

A special form of the radio tower is the *telescopic mast*. These can be erected very quickly. Telescopic masts are used predominantly in setting up temporary radio links for reporting on major news events, and for temporary communications in emergencies. They are also used in tactical military networks. They can save money by needing to withstand high winds only when raised, and as such are widely used in amateur radio.

Telescopic masts consist of two or more concentric sections and come in two principal types:

- Pump-up masts are often used on vehicles, and are raised to their full height pneumatically or hydraulically. They are usually only strong enough to support fairly small antennas.
- Telescopic lattice masts are raised by means of a winch, which may be powered by hand or an electric motor. These tend to cater for greater heights and loads than the pump-up type. When retracted, the whole assembly can sometimes be lowered to a horizontal position by means of a second tiltover winch. This enables antennas to be fitted and adjusted at ground level before winching the mast up.

Balloons and kites

A tethered balloon or a kite can serve as a temporary support. It can carry an antenna or a wire (for VLF, LW or MW) up to an appropriate height. Such an arrangement is used

occasionally by military agencies or radio amateurs. The American broadcasters TV Martí broadcast a television program to Cuba by means of such a balloon.

Other special structures

For two VLF transmitters wire antennas spun across deep valleys are used. The wires are supported by small masts or towers or rock anchors. The same technique was also used for the Criggion VLF transmitter.

For ELF transmitters ground dipole antennas are used. Such structures require no tall masts. They consist of two electrodes buried deep in the ground at least a few dozen kilometres apart. From the transmitter building to the electrodes, overhead feeder lines run. These lines look like power lines of the 10 kV level, and are installed on similar pylons.

Design features

Economic and aesthetic considerations

A radio amateur's do it yourself steel-lattice tower

Felsenegg-Girstel TV-tower

Uetliberg TV-tower

Communications tower, camouflaged as a slim tree

- The cost of a mast or tower is roughly proportional to the square of its height.
- A guyed mast is cheaper to build than a self-supporting tower of equal height.
- A guyed mast needs additional land to accommodate the guys, and is thus best suited to rural locations where land is relatively cheap. An unguyed tower will fit into a much smaller plot.
- A steel lattice tower is cheaper to build than a concrete tower of equal height.
- Two small towers may be less intrusive, visually, than one big one, especially if they look identical.
- Towers look less ugly if they and the antennas mounted on them appear symmetrical.
- Concrete towers can be built with aesthetic design - and they are, especially in Continental Europe. They are sometimes built in prominent places and include observation decks or restaurants.

Masts for HF/shortwave antennas

For transmissions in the shortwave range, there is little to be gained by raising the antenna more than a few wavelengths above ground level. Shortwave transmitters rarely use masts taller than about 100 metres.

Access for riggers

Because masts, towers and the antennas mounted on them require maintenance, access to the whole of the structure is necessary. Small structures are typically accessed with a ladder. Larger structures, which tend to require more frequent maintenance, may have stairs and sometimes a lift, also called a service elevator.

Aircraft warning features

Tall structures in excess of certain legislated heights are often equipped with aircraft warning lamps, usually red, to warn pilots of the structure's existence. In the past, ruggedized and under-run filament lamps were used to maximize the bulb life. Alternatively, neon lamps were used. Nowadays such lamps tend to use LED arrays.

Height requirements vary across states and countries, and may include additional rules such as requiring a white flashing strobe in the daytime and pulsating red fixtures at night. Structures over a certain height may also be required to be painted with contrasting color schemes such as white and orange or white and red to make them more visible against the sky.

Light pollution and nuisance lighting

In some countries where light pollution is a concern, tower heights may be restricted so as to reduce or eliminate the need for aircraft warning lights. For example in the United States the 1996 Telecommunications Act allows local jurisdictions to set maximum heights for towers, such as limiting tower height to below 200 feet and therefore not requiring aircraft illumination under U.S. Federal Communications Commission (FCC) rules. The limit is more commonly set to 190 or 180 feet to allow for masts extending above the tower.

Wind-induced oscillations

One problem with radio masts is the danger of wind-induced oscillations. This is particularly a concern with steel tube construction. One can reduce this by building cylindrical shock-mounts into the construction. One finds such shock-mounts, which look like cylinders thicker than the mast, for example, at the radio masts of DHO38 in Saterland. There are also constructions, which consist of a free-standing tower (usually from reinforced concrete), onto which a guyed radio mast is installed. The best known such construction is the Gerbrandy Tower in Lopik (the Netherlands). Further towers of this building method can be found near Smilde (the Netherlands) and Fernsehturm, Waldenburg, Baden-Württemberg, Germany).

Hazard to birds

Radio, television and cell towers have been documented to pose a hazard to birds. Reports have been issued documenting known bird fatalities and calling for research to find ways to minimize the hazard that communications towers can pose to birds.

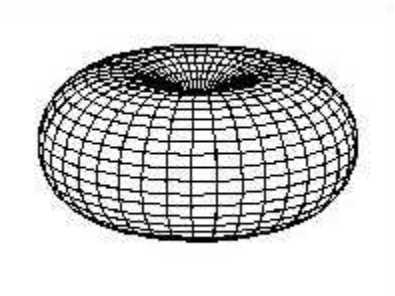
Law

Since June 2010, Telecom operators in the USA can erect new telecom masts or towers as the government has lifted the moratorium, which was earlier placed on the issuance of permits for the construction of telecommunication towers.

Chapter 8

Omnidirectional Antenna & Directional Antenna

Omnidirectional Antenna



Omnidirectional radiation pattern of a vertical dipole antenna. In this graph the antenna is at the center of the "donut". Radial distance from the center represents the power radiated in that direction. The power radiated is maximum in horizontal directions, dropping to zero directly above and below the antenna.

An **omnidirectional antenna** is an antenna which radiates power uniformly in all directions in one plane, with the radiated power decreasing with elevation angle above or below the plane, dropping to zero on the antenna's axis. This radiation pattern is often described as "donut shaped". Note that this is different from an isotropic antenna, in which the gain is uniform in *all* directions ("spherical"). Omnidirectional antennas oriented vertically are widely used for nondirectional antennas on the surface of the Earth because they radiate equally in all horizontal directions, while the power radiated drops off with elevation angle so little radio energy is aimed into the sky or down toward the earth and wasted. Omnidirectional antennas are widely used for radio broadcasting antennas, and in mobile devices that use radio such as cell phones, FM radios, walkie-talkies, Wifi, cordless phones, GPS as well as for base stations that communicate with mobile radios, such as police and taxi dispatchers and aircraft communications.

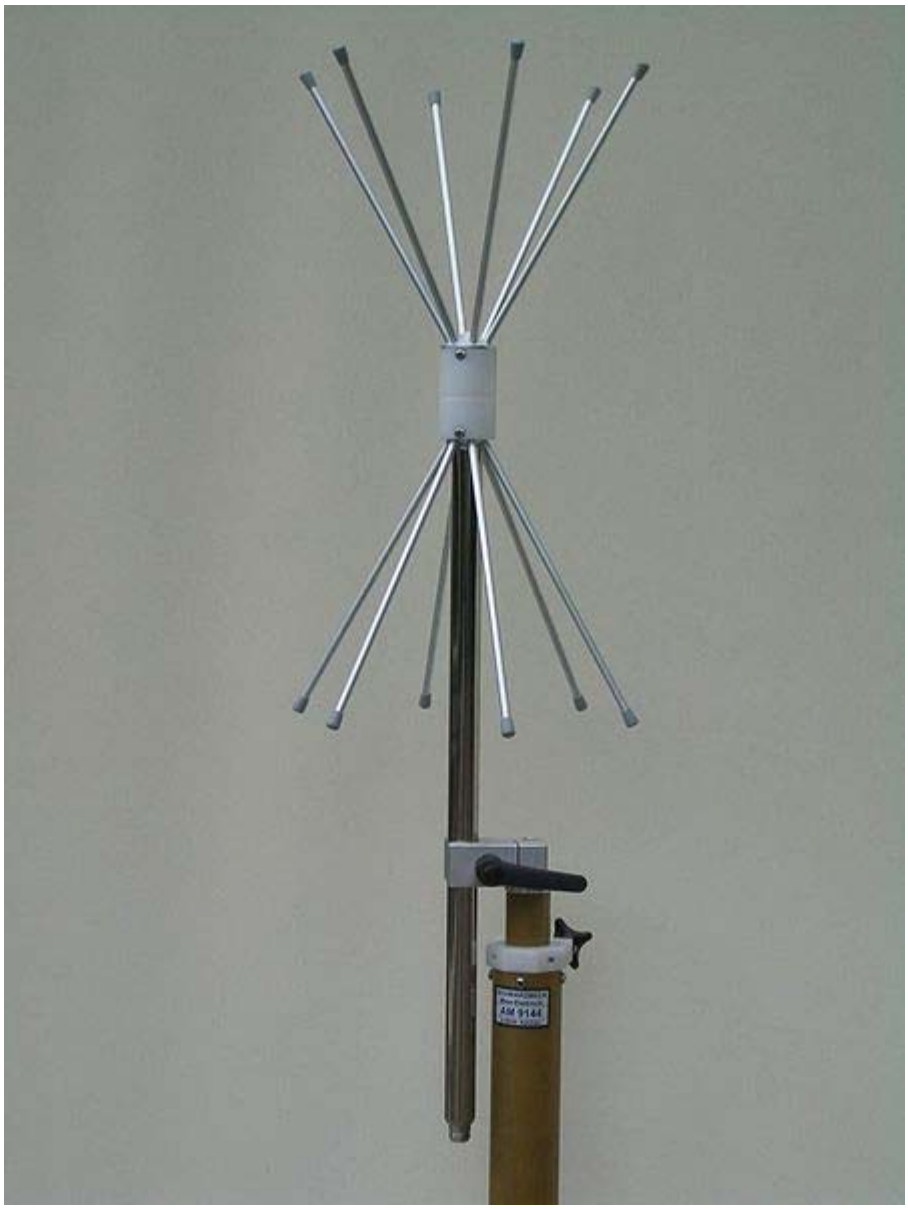
Types

Common types of low gain omnidirectional antennas are the whip antenna, "Rubber Ducky", ground plane antenna, vertically oriented dipole antenna, discone antenna, mast radiator and the horizontal loop antenna (or halo antenna) (Sometimes known colloquially as a 'circular aerial' because of the shape).

Higher gain omnidirectional antennas can also be built. "Higher gain" in this case means that the antenna radiates less energy at higher and lower elevation angles and more in the horizontal directions. High gain omnidirectional antennas are generally realized using collinear dipole arrays. These arrays consist of half-wavelength dipoles with a phase shifting method between each element that ensures the current in each dipole is in phase. The Coaxial Colinear or COCO antenna uses transposed coaxial sections to produce in-phase half-wavelength radiatiors. A Franklin Array uses short U-shaped half-wavelength sections whose radiation cancels in the far-field to bring each half-wavelength dipole section into equal phase.

Types of higher gain omnidirectional antennas are the Coaxial Colinear (COCO) antenna and Omnidirectional Microstrip Antenna (OMA).

Some planar antennas (constructed from printed circuit board) are omnidirectional antennas.



Vertical polarized VHF- UHF biconical antenna 170 – 1100 MHz with omni directional H-plane pattern.

Analysis

Omnidirectional radiation patterns are produced by the simplest practical antennas, monopole and dipole antennas, consisting of one or two straight rod conductors on a common axis. Antenna gain (G) is defined as antenna efficiency (e) multiplied by antenna directivity (D) which is expressed mathematically as: $G = eD$. A useful relationship between omnidirectional radiation pattern directivity (D) in decibels and half-power beamwidth (HPBW) based on the assumption of a $\sin b\theta$ / $b\theta$ pattern shape is:

$$D = 10 \log_{10} \left(\frac{101.5}{HPBW - 0.00272(HPBW)^2} \right) \ dB$$

Directional Antenna

Log-periodic dipole array

A **directional antenna** or **beam antenna** is an antenna which radiates greater power in one or more directions allowing for increased performance on transmit and receive and

reduced interference from unwanted sources. Directional antennas like yagi antennas provide increased performance over dipole antennas when a greater concentration of radiation in a certain direction is desired.

All practical antennas are at least somewhat directional, although usually only the direction in the plane parallel to the earth is considered, and practical antennas can easily be omnidirectional in one plane.

The most common types are the yagi antenna, the log-periodic antenna, and the corner reflector, which are frequently combined and commercially sold as residential TV antennas. Cellular repeaters often make use of external directional antennas to give a far greater signal than can be obtained on a standard cell phone. Satellite Television receivers usually use parabolic antennas.

For long and medium wavelength frequencies, tower arrays are used in most cases as directional antennas.

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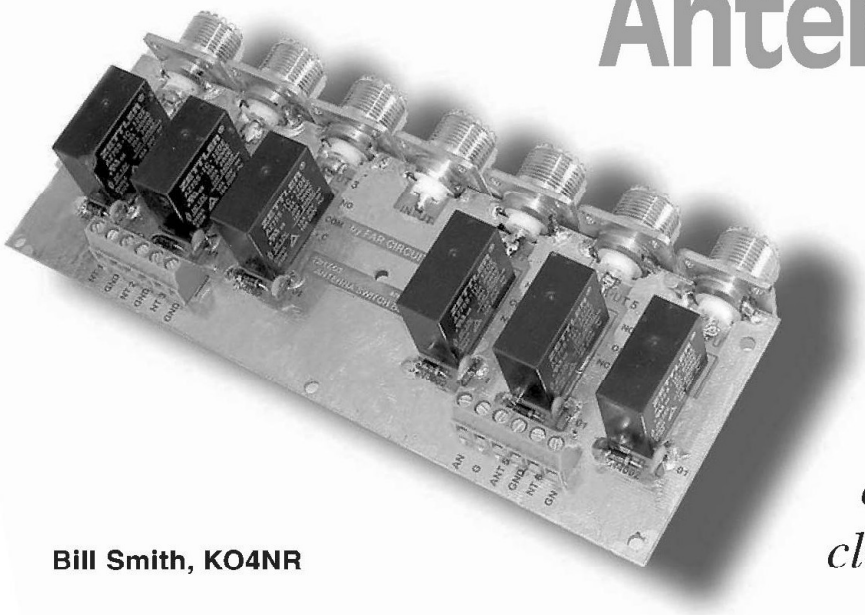
A Low-Cost Remote Antenna Switch



Bill Smith, KO4NR

Here's an easy and inexpensive way to reduce the number of antenna feed lines cluttering up your shack.

How many of us have struggled with getting multiple antenna feed lines into our shacks? I believe it is a common problem that can be easily solved with a suitable remote antenna switch.

My shack is in the basement on the opposite end of the house from the antenna feed lines. My antenna farm consists of a 2 meter beam, an end-fed half-wave sloper and a modified Zepp by NB6Z. I pondered the problem of getting the feed lines for all the antennas into the shack for several days without coming up with a good plan. My dilemma was finding a path to run three or more coaxial feed lines across the top of the basement's suspended ceiling. In addition I would have to win the approval of my wife.

I cautiously began the easement negotiations knowing it would take some time. Negotiations were complicated by the fact that a hole large enough for three or more coaxial feed lines would have to be cut in the house's wood siding. In the end, my wife agreed to a plan that included a small hole in the wood siding and a single coax run across the top of the basement's suspended ceiling. Switching antennas meant frequent trips outside to connect the right feed line to the coax running into the house. I operated this way for over four years, all the while longing for a better way to switch my antennas.

Remote antenna switches were the answer, but they can be expensive. I had seriously considered purchasing one of the many remote antenna switches on the market, but I couldn't decide which one to go with. As if struck by lightning, it came to me: Why not build a switch using one of the many printed circuit board (PCB) power relays on the market? Surely one of the hundreds of PCB power relays available would be suitable for switching antennas.

Relay Selection

The solution to my antenna switching problem presented itself while I was experimenting with PCB relays for use in a linear amplifier's tuned input circuit. The search for a suitable PCB relay for the amplifier revealed interesting innovations in PCB relay design. They now come in small packages, exhibit high dielectric strength and can carry impressive amounts of current.

Although not factory tested for RF use, I found the American Zettler AZ755 series PCB relays to work very well.[1] The AZ755 is rated for 480 W switched power with a resistive load and a maximum switched current of 20 A. Despite the relay's small physical size, the dielectric strength between the contacts and coil is 5 kV RMS, with an impressive 1 kV RMS between the open contacts. This means that the relay is resistant to a flashover that could damage the coil or pit the contacts.

The AZ755 series relays are offered in a wide variety of configurations. I used American Zettler part number AZ755-1C-12DE. This model has a 12 V dc coil, but you can use any of the available coil voltages in the series. The contact style is Form C, which is single-pole double-throw (SPDT). The E suffix indicates that the relay is epoxy sealed. I thought the epoxy seal would provide better protection from dirt and moisture contamination. If you prefer relays that are not epoxy sealed, drop the E from the part number.

American Zettler relays are readily available on-line from **RelayCenter.com**.[2]

Circuit Design and Board

Figure 1 shows the final circuit. I settled on a switch that would handle up to six antennas, enough for my current antenna farm and possible future expansion. The common contacts of the relays (K1-K6) are connected to SO-239 connectors (J1-J6) for the antenna feed lines. The normally open (NO) contacts are all connected to the RF INPUT connector, J7.

[1]Notes appear on page 41.

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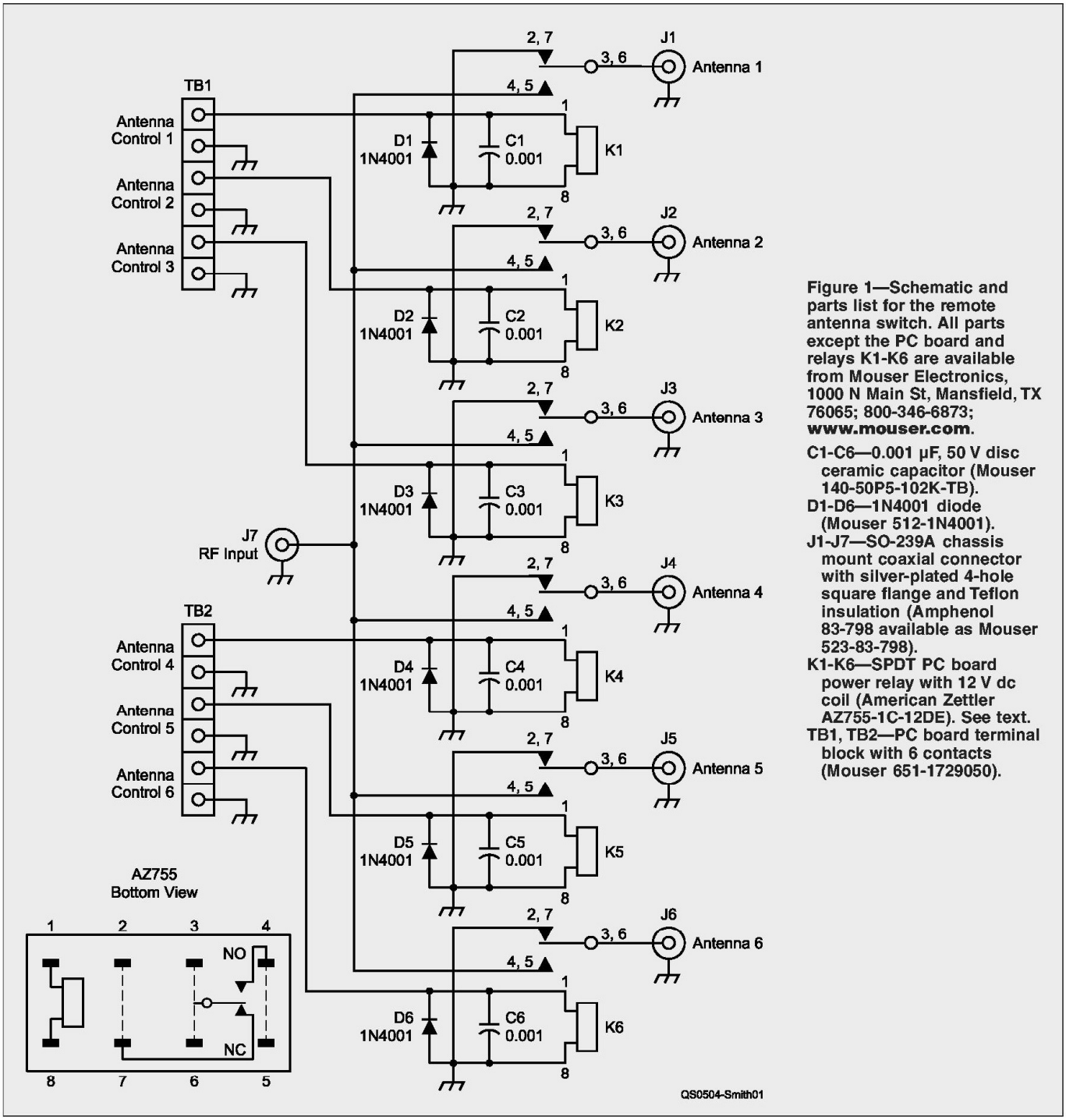

Figure 1—Schematic and parts list for the remote antenna switch. All parts except the PC board and relays K1-K6 are available from Mouser Electronics, 1000 N Main St, Mansfield, TX 76065; 800-346-6873; www.mouser.com.

C1-C6—0.001 µF, 50 V disc ceramic capacitor (Mouser 140-50P5-102K-TB).
D1-D6—1N4001 diode (Mouser 512-1N4001).
J1-J7—SO-239A chassis mount coaxial connector with silver-plated 4-hole square flange and Teflon insulation (Amphenol 83-798 available as Mouser 523-83-798).
K1-K6—SPDT PC board power relay with 12 V dc coil (American Zettler AZ755-1C-12DE). See text.
TB1, TB2—PC board terminal block with 6 contacts (Mouser 651-1729050).

The normally closed (NC) contacts are all connected to ground so that the antennas are grounded when not in use. To select an antenna, apply 12 V dc to the appropriate ANTENNA CONTROL terminal to energize the relay and connect the ANTENNA to the RF INPUT.

To keep stray RF out, 0.001 µF ceramic disc capacitors (C1-C6) are installed across the relay coils. In addition, 1N4001 diodes (D1-D6) are installed across the coils to prevent voltage spikes when the power is removed from the coil.

Once I located a relay that might work and had a circuit in mind, I contacted FAR Circuits to see if an inexpensive circuit board could be produced. FAR Circuits has built several boards for my linear amplifier projects and has a proven history of supplying high quality circuit boards. Fred at FAR Circuits agreed to design the board[3] for me using the AZ755 relays with SO-239 input and output connectors. He suggested mounting the SO-239s directly to the board to eliminate wiring and minimize SWR problems. The finished board with all components mounted is shown in the title photo. The RF INPUT connector is in the center, with three ANTENNA connectors on each side. The control cable connects to the two terminal blocks.

Assembly Notes

The design is simple and assembly doesn't require special tools. Far too often, projects require sophisticated test equipment and a degree of expertise I don't possess. This project can easily be completed by anyone with basic soldering skills.

In addition to the PC board and parts, you'll need a suitable enclosure to keep

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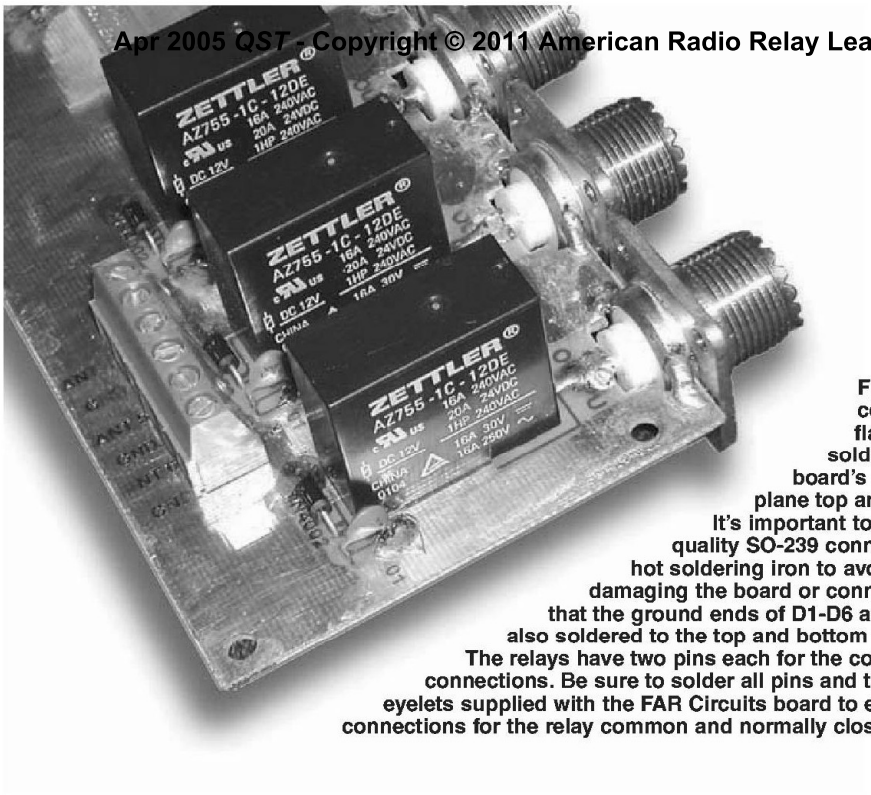

Figure 2—The connector flanges are soldered to the board's ground plane top and bottom. It's important to use good quality SO-239 connectors and a hot soldering iron to avoid damaging the board or connectors. Note that the ground ends of D1-D6 and C1-C6 are also soldered to the top and bottom of the board. The relays have two pins each for the contact connections. Be sure to solder all pins and to install eyelets supplied with the FAR Circuits board to ensure good connections for the relay common and normally closed pins.

the board dry and pests away. I installed mine in a plastic children's lunch box that hangs under my deck. Initially this enclosure was just for testing purposes, but it has worked out so well that I may not change it.

The most difficult part of the project is drilling the holes in the enclosure for the SO-239 connectors and getting everything to line up. You may find it easier to install the connectors in the enclosure first, and then solder them to the board. (Use the board to mark the center line and location of the connectors on the enclosure.) After the connectors are tacked in place, remove the screws holding the SO-239s to the enclosure and remove the total assembly. This ensures a perfect fit when reassembling.

Make sure the SO-239s are all the same type and brand to ensure a uniform fit. The board was designed around the Amphenol connectors recommended in the parts list. They have silver-plated center pins and flanges, and Teflon insulation. The silver plating makes the connectors easier to solder than nickel-plated connectors, and the Teflon insulation is much less prone to melting than the plastic often found on inexpensive connectors. Of course you can use other SO-239 connectors, but you may have to modify the board for a good fit.

Use a *hot* soldering iron when soldering the flange of the SO-239 to the board's ground plane as shown in Figure 2. I used a Sears Craftsman Dual 230/150 W iron and it worked great. Be sure to solder top and bottom.

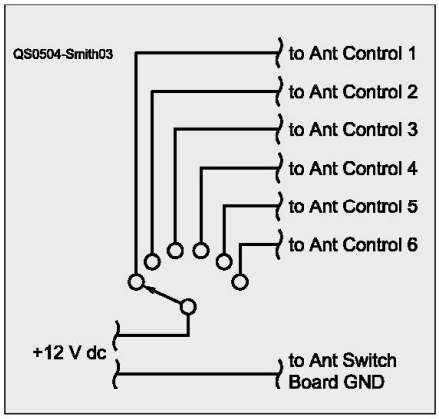

Figure 3—I used a simple 12 V supply and rotary switch in my shack to select antennas.

The PC board is double-sided, and FAR Circuits supplies eyelets with the board to use in the larger holes for the common and normally closed relay pins. They provide for a better connection to the component side of the board. In addition, soldering a short length of no. 14 bare copper wire into the center pin of each SO-239 will give you a better connection to the circuit trace on the board.

Powering It Up

To control the switch, I use a 12 V dc power supply in the shack and a small ceramic rotary switch as shown in Figure 3. The switch simply sends 12 V to energize one of the relay coils and select the desired antenna. When 12 V is removed, the normally closed relay contacts connect all feed lines to ground. Connect the PC board's ground plane to your outside lightning protection ground system.

I ran an 8-conductor control cable into the shack using the same route taken when I installed the coax. My wife agreed to it when I explained the cable would eliminate the need to run additional coax into the house. My control cable has #16 conductors and is something I had in my parts collection. There is a chart at **www.altronix.com/html/an101.htm** that you can use to help you determine the right wire size for your installation. Long runs of wire may require larger conductors to prevent unacceptable voltage drops at the relay coils, but the relays will work over a fairly wide range of coil voltages so it's not critical.

Test Results

Once the switch was completed, I performed several tests to see how it worked. The testing was done with my Kenwood TS-850S transceiver, RadioShack MTA-20 digital power meter and Swan Mark 1 linear amplifier. The board was not installed in an enclosure for the initial testing. It was lying on top of a plastic box.

Test 1. For the initial test, I connected the TS-850S (with internal antenna tuner off) to the RF INPUT jack on the antenna switch board. Then I connected one of the antenna jacks on the switch board to the RadioShack power meter and then to a dummy antenna. From 160-10 meters, the internal meter in the TS-850S indicated that the SWR was about 1.2:1, and the external power meter indicated no loss of power through the relay board. My TS-850S is sensitive to SWR and has always folded the power back when transmitting into a high SWR. That didn't happen, so I concluded the SWR with the board in the line must be low.

Test 2. Next I added the Swan Mark 1 amplifier between the transceiver and antenna switch. I measured the power with the digital power meter before and after the antenna switch board and could detect no difference. I operated the amplifier in the CW mode at 1-1.3 kW. The relay got a little warm but no more than I would expect from an energized coil, and it seemed to handle the power just fine. SWR with the switch in line remained low on all bands.

Test 3. For the final test, I borrowed a friend's MFJ-259B antenna analyzer to use as a reality check. For this test, I

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placed the board under my deck where it would be in normal use, connected it to a dummy antenna, and ran about 70 feet of RG-8X to the antenna analyzer. I then used the MFJ-259B to measure the SWR, reactance (X_s) and resistance (R_s) on each band from 1.8-144 MHz. The results are shown in Table 1.

ARRL Lab Testing. The ARRL Lab had an opportunity to test the completed antenna switch board as well. Insertion loss measured <0.1 dB for 2-50 MHz (for all ports to common). SWR measured 1.1:1 or less from 2-28 MHz, 1.2:1 or less on 50 MHz. Isolation was >60 dB for 2-28 MHz, except for the two innermost ports, which were 50 dB at 28 MHz. Worst-case isolation on 50 MHz was 45 dB.

The various tests showed that although they are not designed for RF, the relays seem to perform well. The board exhibits low SWR, low insertion loss and good isolation over a wide frequency range. It handles 1-1.3 kW during normal intermittent SSB and CW operation (I did not try it with a key-down mode like RTTY).

Summary

This project offers a rewarding solution to managing the number of feed lines snaking through your house. Parts are readily available, and the FAR Circuits PC board makes construction straightforward. If homebrewing is not your cup of tea, don't despair. There are several good commercial antenna switches available from the advertisers in *QST*.

Table 1
MFJ-259B Test Results

Freq (MHz)	SWR	X_s (Ω)	R_s (Ω)
1.8	1.1:1	0	63
2.0	1.1:1	0	61
3.5	1.1:1	5	53
4.0	1.0:1	0	56
7.0	1.0:1	0	57
7.3	1.0:1	0	58
10.1	1.2:1	9	51
10.15	1.1:1	7	49
14.0	1.1:1	7	50
14.35	1.1:1	4	53
18.068	1.1:1	2	50
18.168	1.0:1	0	55
21.0	1.2:1	9	48
21.45	1.1:1	8	52
24.89	1.4:1	5	53
24.99	1.1:1	3	53
28.0	1.1:1	6	50
29.7	1.0:1	0	61
50	1.0:1	0	57
54	1.0:1	0	61
144	1.2:1	11	47
148	1.2:1	9	47

Notes

[1]You can find more information at American Zettler's Web site, **www.americanzettler.com**. A data sheet for the AZ755 series relays may be downloaded from **www.azettler.com/pdfs/az755.pdf**.

[2]**www.relaycenter.com**. At publication time, the AZ755-1C-12DE relays used in this project were $2.25 each plus shipping and handling, with a minimum order of $25.

[3]The circuit board is available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118 for $11 plus $2 shipping and handling. For ordering information, see the FAR Circuits Web site at **www.farcircuits.net** and look under the Repeater Controller & Station Accessory heading.

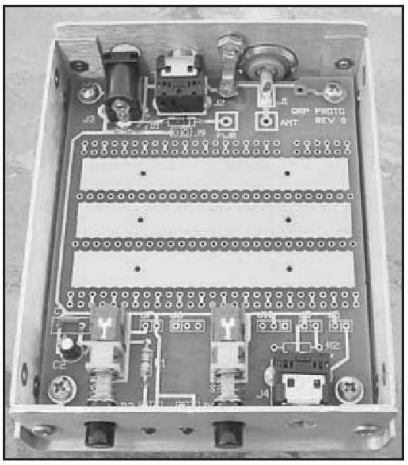
Bill Smith, KO4NR, enjoys repairing and modifying vintage linear amplifiers, including a Swan Mark 1 and Heath Warrior. He usually operates 17 and 40 meter SSB, and especially likes chatting with other hams who share his passion for building and working on equipment. Employed as maintenance manager at a gas fired energy plant, Bill can be reached at 244 Cameron St, Manchester, NH 03103, or by e-mail at **ko4nrbs@yahoo.com**.

New Products

QRP PROTOTYPING KIT

◊ The QRP Prototyping Kit has been added to the Alden McDuffie kit product line. This kit is designed to save time for those constructing a project. The kit includes case, printed circuit board (PCB) with "dead bug" construction area and pads with layout for SO-239 coaxial connectors, two 3.5 mm audio jacks and two alternate-action switches. All those parts are included. The printed circuit board is designed to fit in their 3.2×4×1.25 inch project case.

A step-by-step manual is designed to assist in building the kit and planning the project. Included are the case, mechanical drawings, schematic and parts list for the PCB kit. Price: fully assembled with case, $54.95; full kit, $42.95; PCB kit without case, $29.95. For more information, contact Alden McDuffie, PO Box 3636, Lawrence, KS 66046; tel 785-766-0404; **www.aldenmcduffie.com**.



PACKET TERMINAL SOFTWARE

◊ *PacTerm 3 for Windows* and *PkTerm 3 for Windows* are offered in preview version by Creative Services Software.

This new release includes a new user interface in a single window, a built-in generic logging program, the MT63 soundcard mode, PSK 62.5 mode, TCP/IP support and more.

New sound card modes just plug in. This means a ham can create a dynamic linked library (DLL) file for any mode, put that DLL into the *PacTerm* or *PkTerm* folder and that new mode appears on the HF mode menu of *PacTerm* or *PkTerm 3*.

The software includes sample skeleton source code to allow a user to create their own modem (HH_DUMB.DLL and source code). Also included is the source code for the HH_MT63.DLL.

Neither program requires a TNC. There is a soundcard only mode that will work with the RIGblaster, MFJ units, Timewave's soon to be released Hamhub and homebrew interfaces.

Price: *PacTerm for Windows and PkTerm for Windows*, $99.95; upgrades from earlier versions, $49.95. Both programs include a 30 day full featured demo version, so hams and MARS users can try them out before purchase. For more information or to order, see **www.cssincorp.com**.

PRACTICAL ANTENNA DESIGN

140-150 MHZ
VHF TRANSCEIVERS

Online Edition

PRACTICAL ANTENNA DESIGN

140-150 MHZ
VHF TRANSCEIVERS

Online Edition

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DEDICATION

This book is sincerely dedicated to the memories of
Ricardo, Mimi, Nida and Poloy
who
through their examples
taught me how to be responsible, strong and steadfast
for what one aspires for.
To them I give my deepest respect.
They died but their ideals will survive...

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INTRODUCTION

This book is one of a series designed to help anyone who wants to construct antennas for radio transceivers but who has only a basic knowledge in radio-communications technology. The approach applied in this book is similar to the do-it-yourself methods of trade books and theories are kept to a minimum. Detailed illustrations are extensively used throughout the whole process of antenna construction to simplify the otherwise difficult to comprehend technical jargon.

The antenna designs presented here are specifically cut to the dimensions necessary for proper operation in 140-150 MHz VHF band. Each chapter deals with a particular design and an extra chapter at the last part is added to help the constructor in converting the given antenna dimensions for other frequencies. However, the formulas for conversion give only a generalized information and much of the fine-tuning of the new dimensions is left to the actual experimentation of the constructor. A highly detailed no-guessing antenna dimensions for other frequency bands are described in other books in this series written by the author.

The choice of a certain design for a particular application is left to the decision of the constructor. In selecting a design, certain factors like portability, ruggedness, compactness, signal gain versus size, weight, wind loading and availability of materials must be taken into account to realize an optimum performance from a particular antenna.

The author assumes that the interested constructor has already some experience in basic construction techniques related to radiocommunications equipment installation like soldering VHF connectors to coaxial cables, making a pig tail, cutting aluminum tubes and using an SWR meter. Obviously a knowledge in operating a VHF transceiver is the most important.

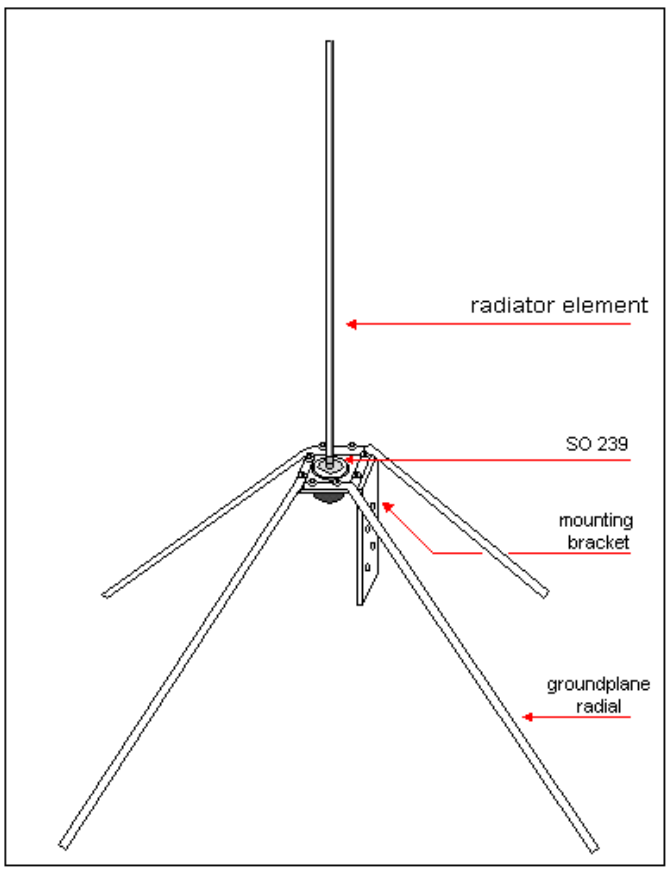
Here is one rule of a thumb in installing VHF antennas: If you use an RG-58/U coaxial cable to feed the antenna, do not use more than 20 meters or 60 feet long. More than this length, much of the signal (almost half) is lost in the cable and will substantially degrade your antenna's performance. If it is unavoidable to extend this length, use the larger RG-8/U cable instead. Although this cable is about four times more expensive than RG-58/U cable, this is the only way you can avoid signal losses in the cable.

It is the author's hope that this book will provide adequate information to anyone wishing to build his own antennas for VHF tranceivers.

1 GROUNDPLANE ANTENNA

Model FA-2

Reliable communications in radio systems depends upon the over-all effectiveness of both the base station and mobile unit antennas. The radiation pattern of the transmitted signal is extremely important since it must be transmitted and received in densely populated areas as well as over long distances. If you are situated in the center of a town or a city, omnidirectional pattern is best suited for you. Omni-pattern is also the best choice when you do not know the exact direction or location of the station you are communicating with. Directive pattern is practical only if you know exactly which direction must the signal be beamed to, in order to maximize the transfer of RF energy. However, antennas with directive patterns are more complex in design and will be discussed in later chapters.



Generally, antennas for VHF bands are mounted as high off the ground as practical to overcome the limitations of the so called line-of-sight transmission and reception. An artificial ground must then be used since the antenna is well above the ground in this case. This is not a problem in automobiles since this artificial ground is provided by either the metal roof or body of the car. For tower installations however, some means must be provided to simulate this artificial ground. This is accomplished by the groundplane radials which are usually made of thin metal rods or tubes each cut to quarterwavelength long and mounted at the base of the antenna. The rods sometimes bend downward at an angle of about 45 degrees below the horizontal. This angle is important to maintain the correct impedance match of the system.

The ease of construction and low cost of a groundplane antenna makes it an ideal choice for VHF operators. The unit described in this chapter uses bronze rods for the radiating element because of their availability and a bronze rod is the easiest to connect to the center pin of the coaxial connector.
The groundplane radials are made of cheaper aluminum tubes. Obviously, the antenna is not easy to disassemble once completed so its use is commonly confined to fixed installations requiring little maintenance.

The operational frequency bandwidth of FA-2 is from 140 MHz up to 150 MHz exhibiting an SWR response of less than 1.5:1 over the entire bandwidth. It has a gain of 1 dB (unity gain) compared to a real dipole. Its signal pattern is omni-directional.

Figure 1.1 Groundplane antenna Model FA-2.

Materials List

Quantity	Specification/Description	Dimensions
4	Aluminum Tubes	3/8 id x 20" each
1	Brass Rod - the brass rod for acetylene welding is recommended	1/8" diameter
1	SO-239 VHF female connector without flange	
8	Stove bolts - brass or stainless	1/8" x 3/4"
8	Lockwashers - brass,stainless or GI	1/8" id
8	Hex nuts - brass, stainless or GI	1/8" id
1	Aluminum plate gauge 14 or 16	2" x 6"
2	U-bolts with accompanying hex nuts and lockwashers	

*id - inside diameter

Construction

First of all construct the antenna mount. It is made from a 1/8'' thick aluminum plate cut to 2'' x 6''. Drill a hole in the plate big enough for the SO-239 VHF connector to insert into (about 5/8'' or 15.8 mm). Drill the hole at the point about 1'' away from one end (see Figure 1.2).

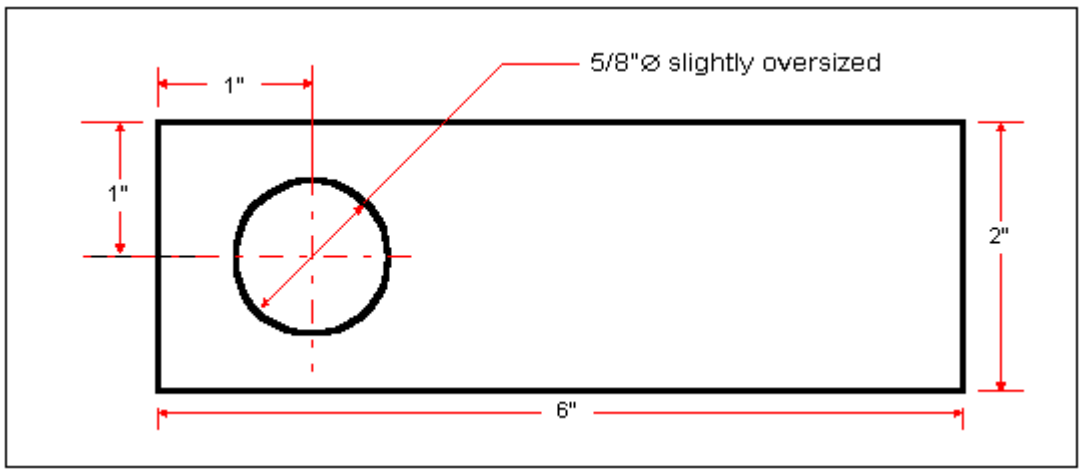


Figure 1.2 Antenna mount and hole Dimensions.

Next drill four holes at the other end of the plate following Figure 1.3 for the proper dimensions. Make sure that the distance between one pair of holes perpendicular to the length of the metal sheet must be the same with the distance of both ends of the U-bolt that will be inserted into it.

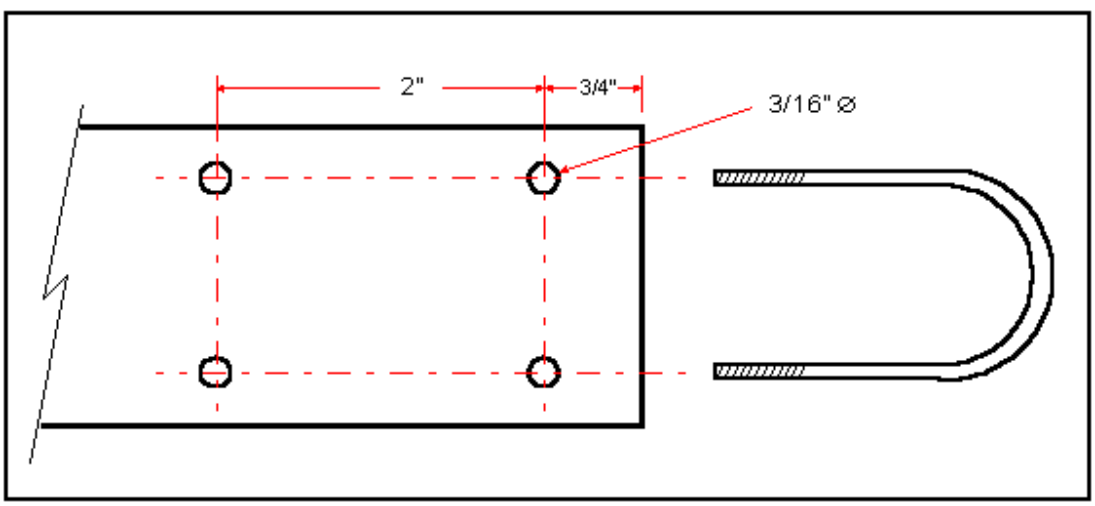


Figure 1.3 Hole dimensions for the U-bolts.

Next, drill eight holes (1/8" diameter) around the large hole following Figure 1.4 for the proper dimensions.

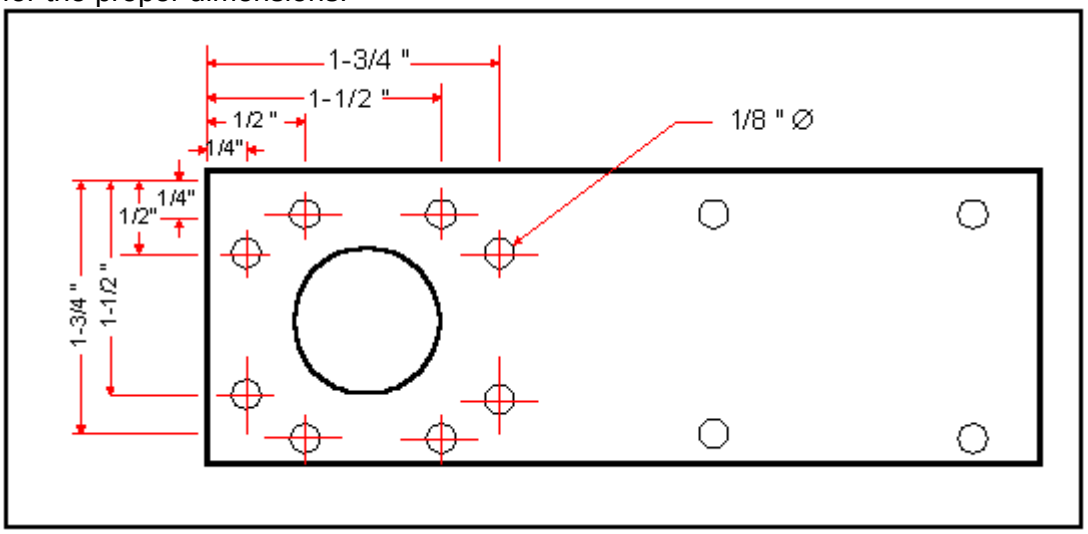


Figure 1.4 Hole dimensions for the radial elements around large hole.

Bend the aluminum plate down to a 90° angle (see Figure 1.5). Follow the illustration for the exact point to bend.

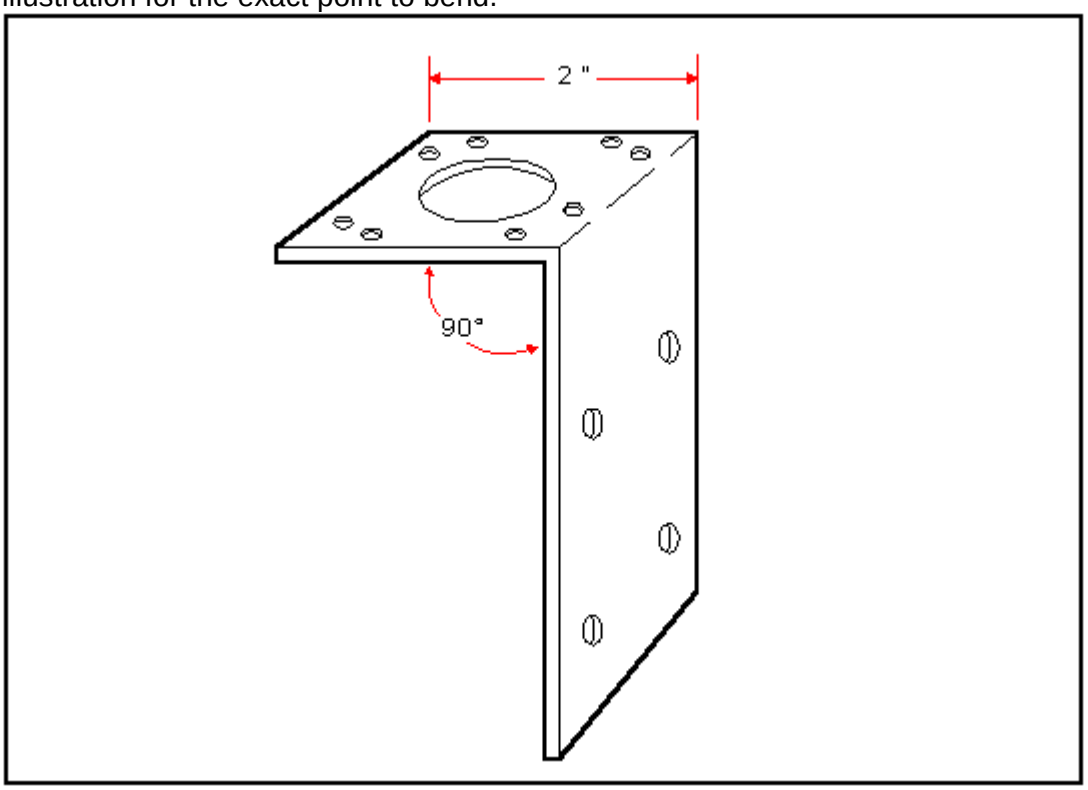


Figure 1.5 Bending the aluminum mounting plate.

Insert the SO-239 VHF connector facing downwards into the mounting plate and fix it permanently with its nut (see Figure 1.6). Discard the grounding ring/lug.

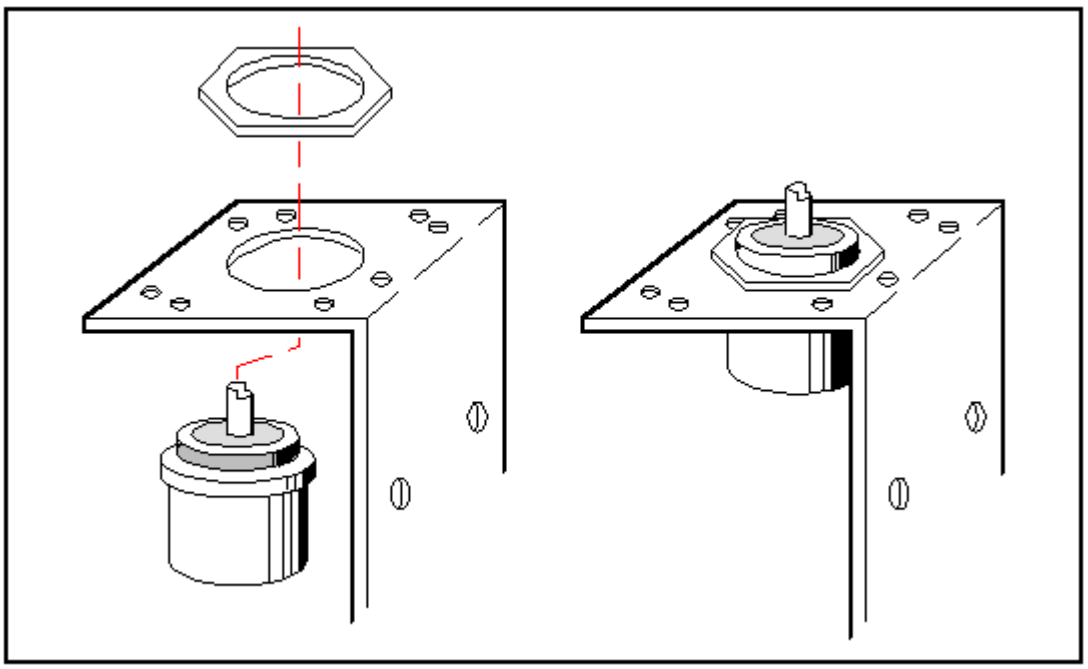


Figure 1.6 Mounting the SO-239 into the plate.

Cut the brass rod to a length of 19" (48.26 cm) and insert one of its end into the center pin of the SO-239 connector (see Figure 1.7). The brass rod may or may not fit into the center pin immediately, so you may need to file away a small portion at the end of the rod to reduce it to a smaller diameter.

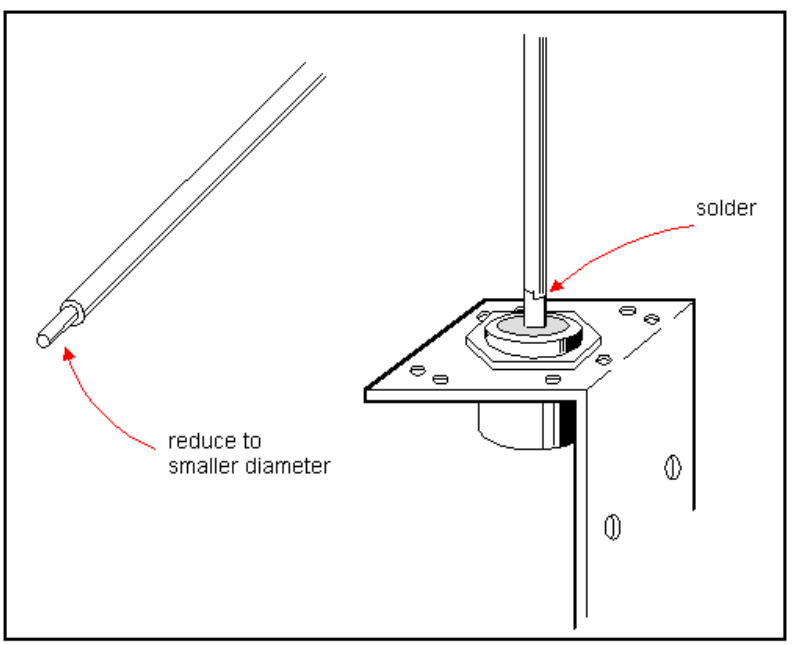


Figure 1.7 Preparing one end of the brass rod to fit inside the SO-239.

Cut four aluminum tubes to a length of 20" each and drill two holes (1/8 " diameter) at one end (see Figure 1.8).

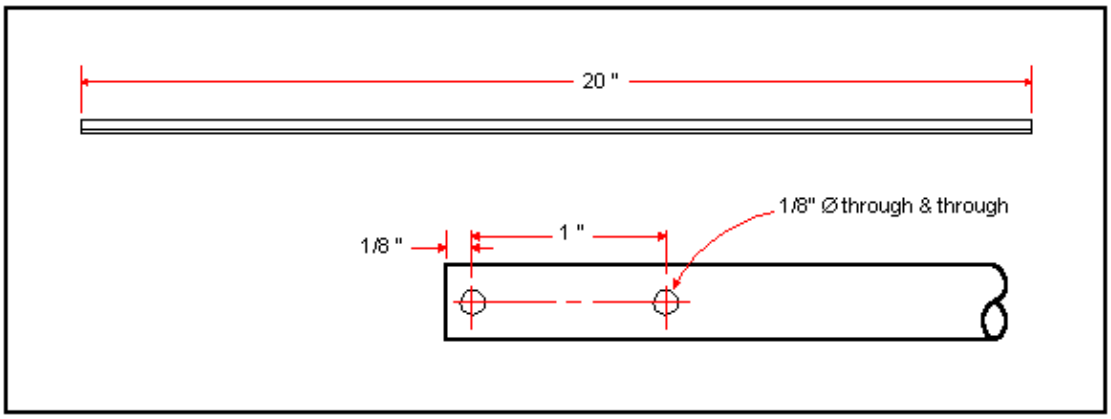


Figure 1.8 Preparing the tubes.

Bend the aluminum tubes to a 45 degree angle at the point 1 inch away for the end with two holes. The direction of the bend must be parallel with the axis of the drilled hole (see Figure 1.9).

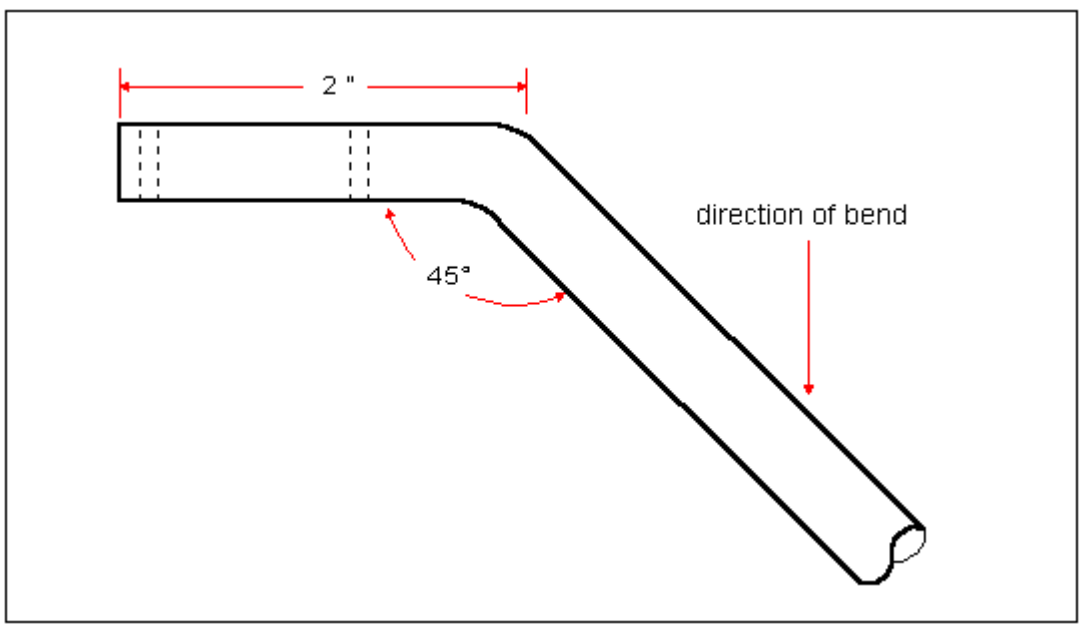


Figure 1.9 Bending the tubes.

Mount the four aluminum tubes into the angled plate by bolting each element with 1/8" x 3/4" stove bolts (see Figure 1.10). The stove bolts must be made of rust resistant material such as stainless steel, brass or GI.

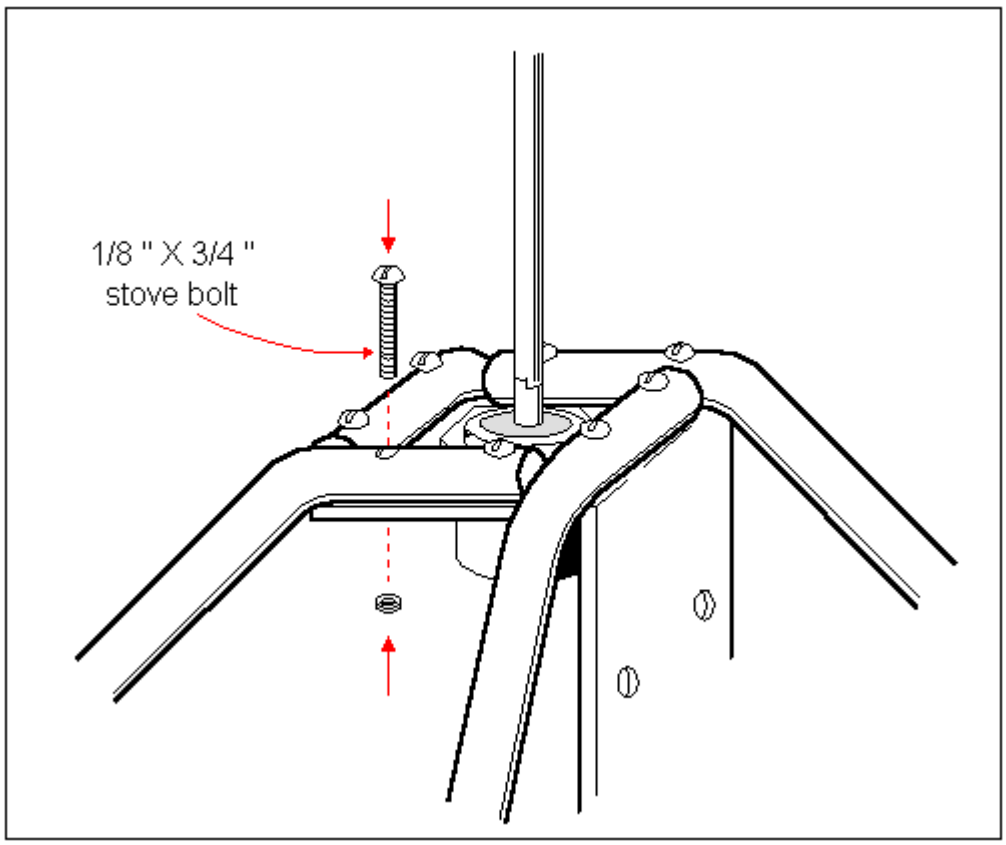


Figure 1.10 Mounting the tubes on the metal plate.

Finally , you can mount the antenna to the mast using the two U-bolts .

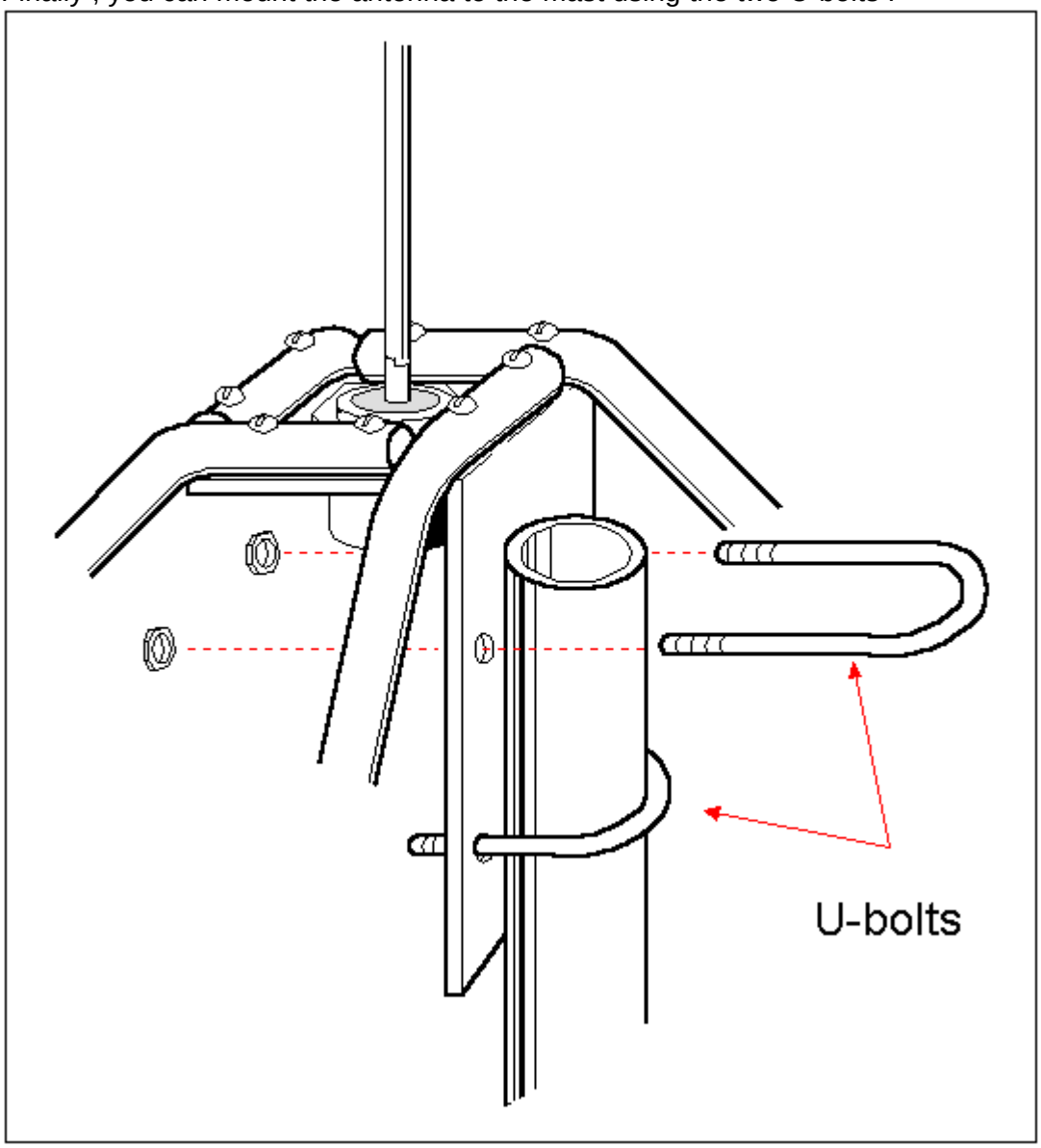
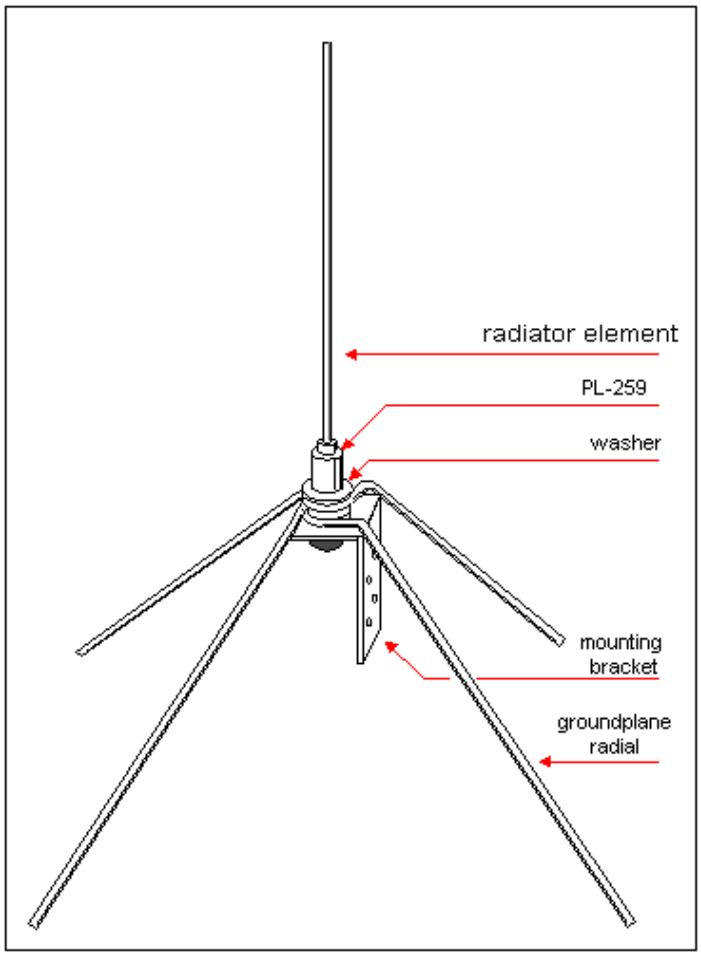


Figure 1.11 Mounting the antenna to the mast.

2 GROUNDPLANE ANTENNA

Model FQ-2



The antenna model FQ-2 is a development from the basic configuration of a groundplane. This unit features quick-detach elements to facilitate for easy and fast disassembly of the antenna. The total size of the antenna is much reduced when disassembled and becomes convenient to carry in transport. It is also a lot easier and faster to construct compared to the FA-2 design.

This particular version of the groundplane was evolved in an emergency situation where there were very few tools available. The place was aboard a fishing boat and there was no drilling tool around, so a groundplane design was created which did not require drilling of holes.

If you plan to use a groundplane in mobile operations, then this design is recommended. It can be easily inserted inside your backpack while you are travelling. Assembly or disassembly takes only a couple of minutes. The antenna elements are made of durable bronze materials so it can survive the stresses caused by the regular mounting and dismounting of the antenna. If you have accidentally bent an element, just straighten it and it is again functional. A slight bend or kink in the elements has no negative effects on the performance of the antenna. It is so durable that you have to intentionally cut it to pieces to destroy it. Experience has proven its reliability in the rugged life of mobile operations.

The electrical characteristics of this antenna are the same with those of model FA-2. The only difference between the two models is the mechanical construction and type of materials used.

Materials List		
Quantity	**Specification/Description**	**Dimensions**
5	Brass rods - the brass rod used for acetylene welding is recommended	1/8" diameter
2	PL-259 VHF male connectors	
1	PL-258 VHF straight connector	
1	Aluminum plate gauge 14 or 16	2" x 6"
2	U-bolts with accompanying hex nuts and lockwashers	
1	Plain washer GI or stainless steel	
1	short length of coaxial cable	2" long

Construction

Reduce one end of a brass rod to a smaller diameter enough to be inserted into the center pin of the PL-259 connector. File also a notch at its end as shown in Figure 2.1b. Insert the rod into the PL-259 and solder it to the center pin (see Figure 2.1c). File away any excess solder that is bulging out of the center pin.

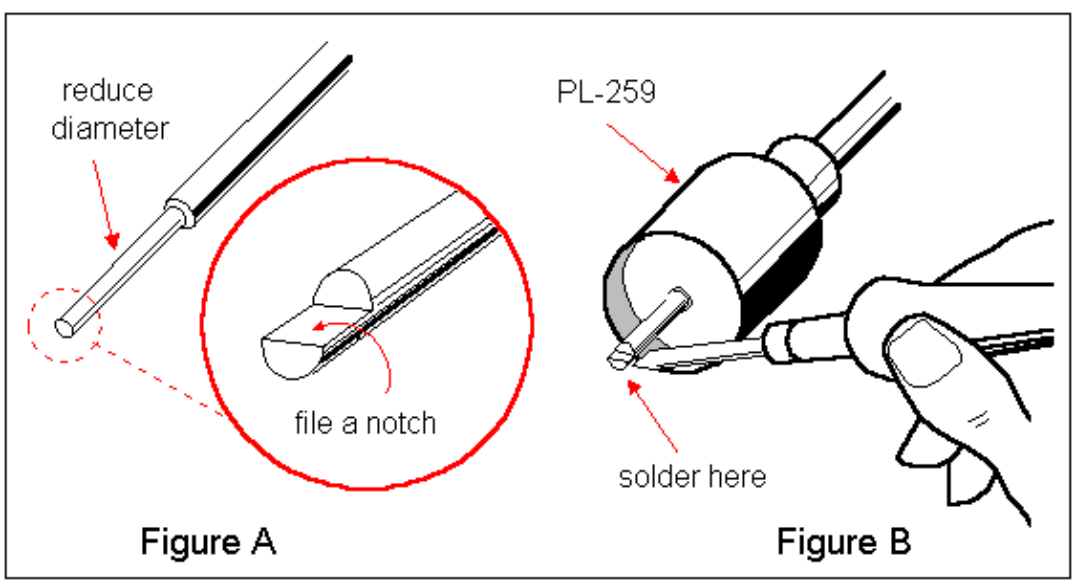


Figure 2.2 Preparing the end of a brass rod and soldering it to PL-259.

After soldering the rod into the PL-259 cut it to a length of 19" (48.26 cm) following Figure 2.3.

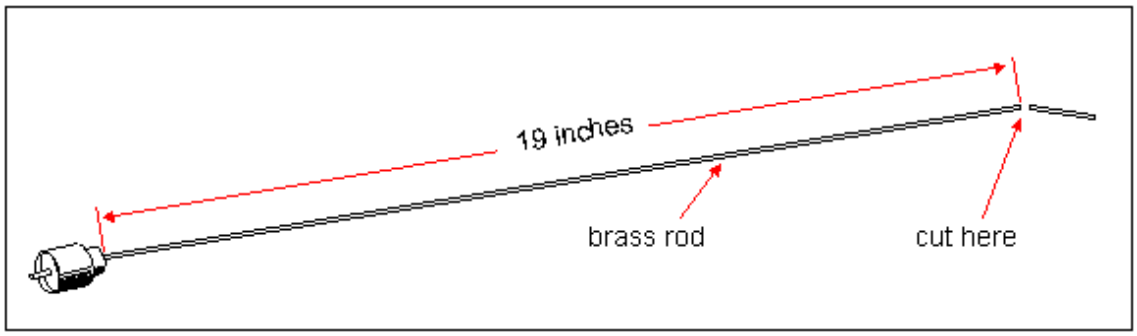


Figure 2.3 Trimming the rod to its proper length.

Cut a small piece of coaxial cable (about 2 inches) and remove its inner conductor and braid (shield). You will only need the vinyl outer jacket. Insert the vinyl jacket into the rod all the way inside the PL-259 (see Figure 2.4). Cut any protruding portion of the jacket. The vinyl jacket serves as an insulator between the brass radiator rod and the body of PL-259.

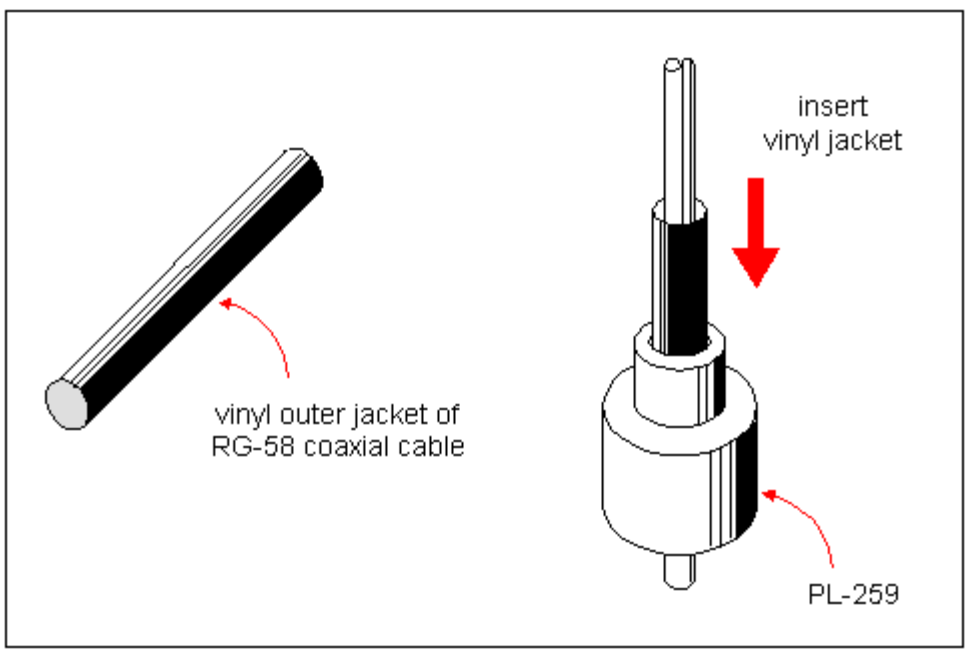


Figure 2.4 Inserting the vinyl insulator into the PL-259.

Prepare a small amount of epoxy glue and place it over the protruding portion of the vinyl insulator. The epoxy glue must cover the gap between the rod and the PL-259 to avoid the seepage of rainwater inside the connector. (Figure 2.5).

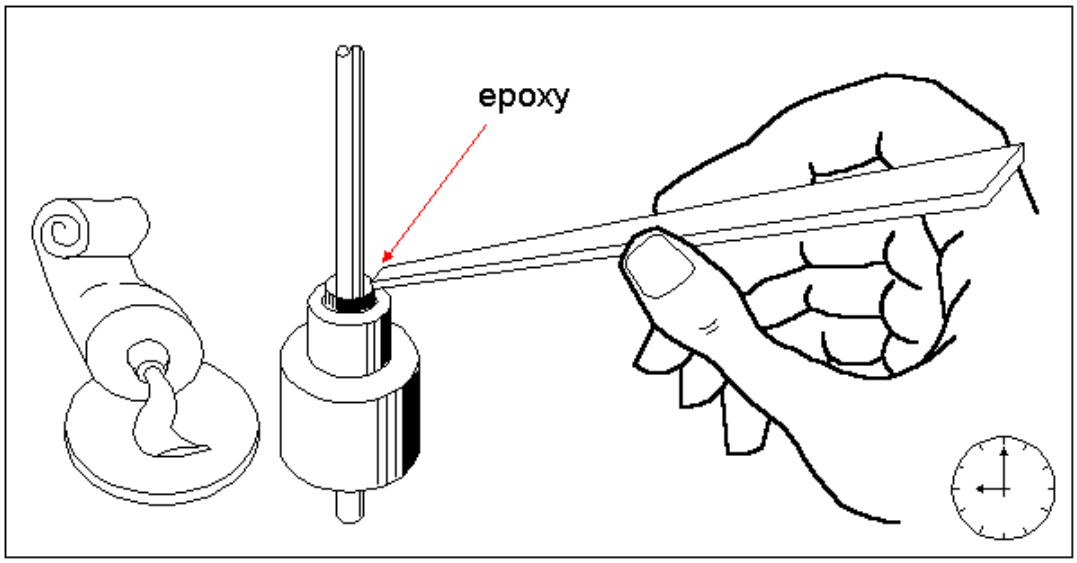


Figure 2.5 Sealing the gap with epoxy glue.

Next prepare the radial elements. Bend one end of each brass rod to an eye-hook shape as shown in the following illustration (Figure 2.6). The diameter of the eyehook form must be dimensioned in such a way that the straight connector can be easily inserted into or pulled out of it.

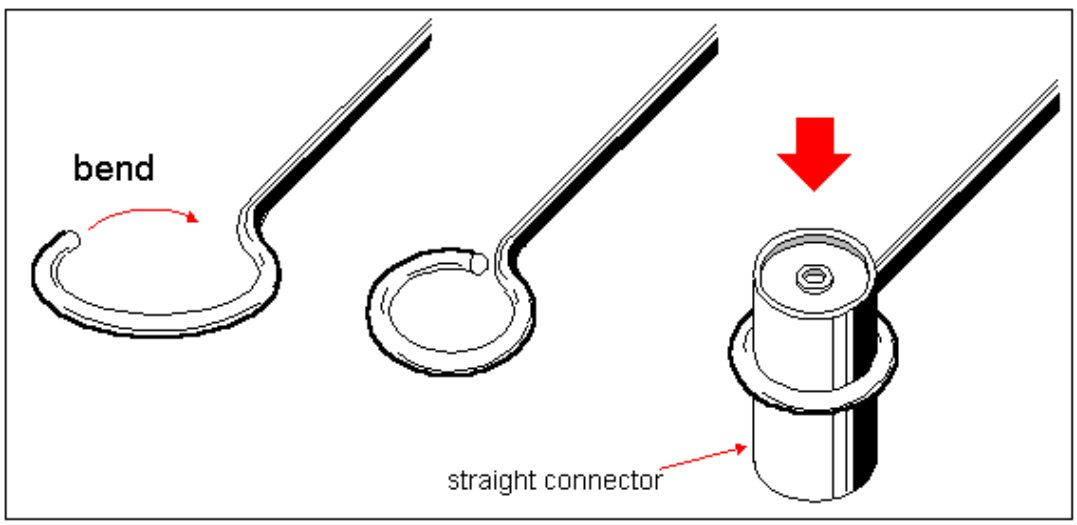


Figure 2.6 Shaping one end of the brass rods to an eyehook form.

After bending one end of all four brass rods into the necessary shape, measure 19 inches from the point where the rod starts to bend into the eyehook form. Mark the measured point at the other end and cut the brass rod at this point (see Figure 2.7).

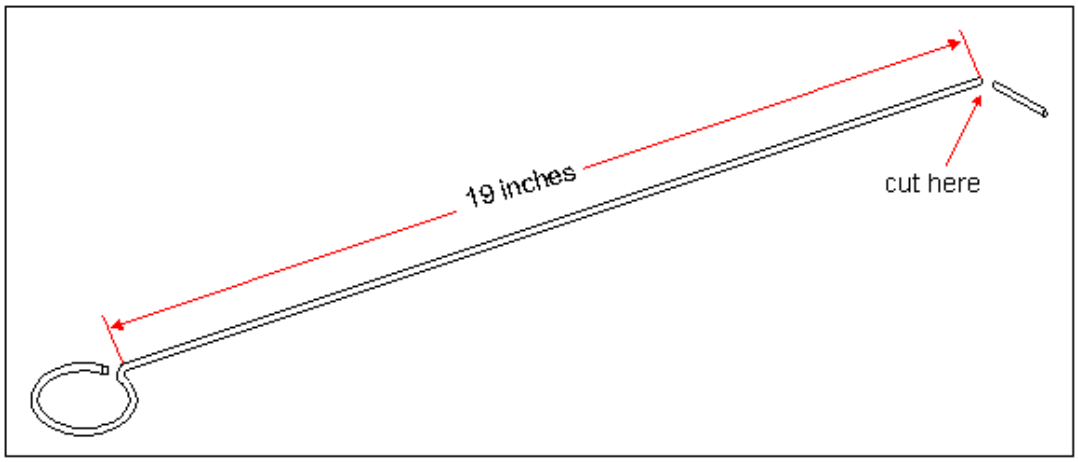


Figure 2.7 Trimming the radial rod to its proper length.

Next, bend the brass rods to a 45 degree angle (see Figure 2.8). Bend the rods at the point 1-1/4 inches away from the center of the eyehook form. The direction of the bend must be perpendicular to the plane of the eyehook end.

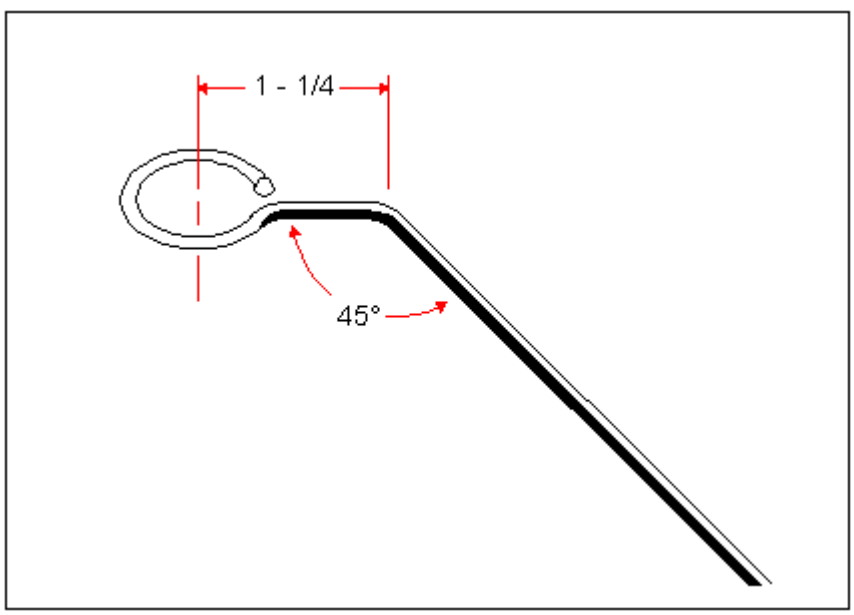


Figure 2.8 Bending the brass radial rod.

Solder the RG-58/U coaxial cable to the remaining PL-259 following the illustrated steps (see Figures 2.9 and Figure 2.10).

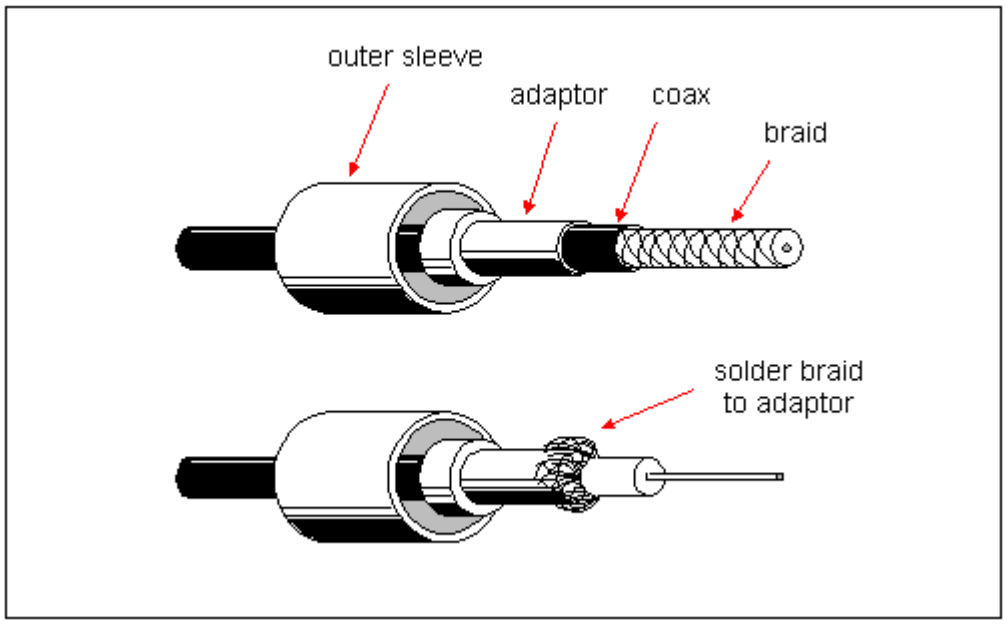


Figure 2.9 Connecting the coaxial cable to the PL-259.

IMPORTANT:
Check the coaxial cable for a possible short after soldering it to the PL-259 connector.

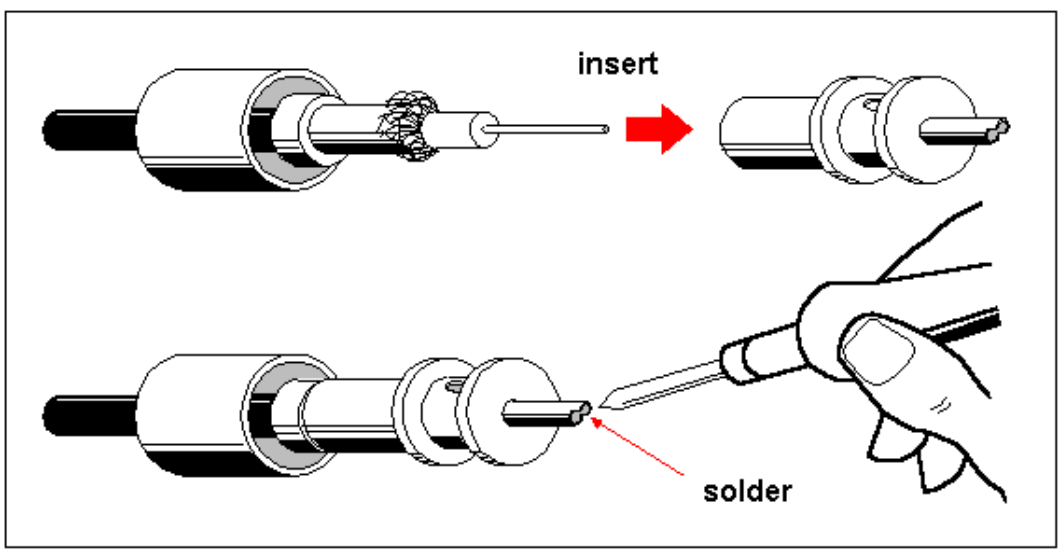


Figure 2.10 Assembling the PL-259.

The mounting bracket for Model FQ-2 is similar to that used for Model FQ-2. The only difference between the two is that the eight small holes around the 5/8" size hole are absent in the bracket for Model FQ-2 (see Figure 2.11).

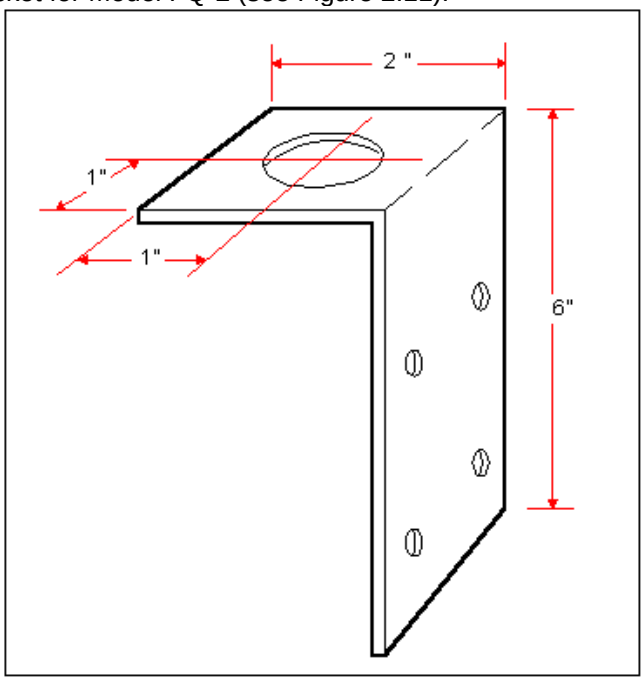


Figure 2.11 Mounting bracket for Model FQ-2.

Assembly

Attach the radiator element into the straight connector. Next, attach the plain washer (5/8" diameter) into the straight connector (see Figure 2.12).

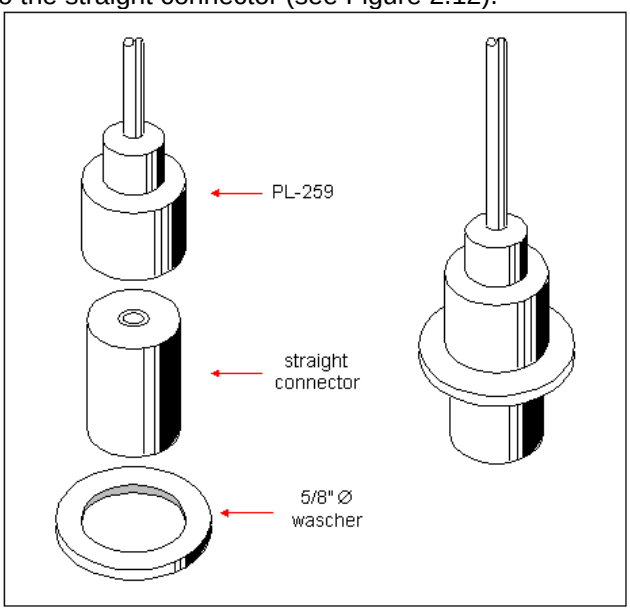


Figure 2.12 Assembling the radiator portion.

Insert the straight connector into the eyehook ends of the brass radial elements. The other ends of the elements must be slooping downwards (see Figure 2.13).

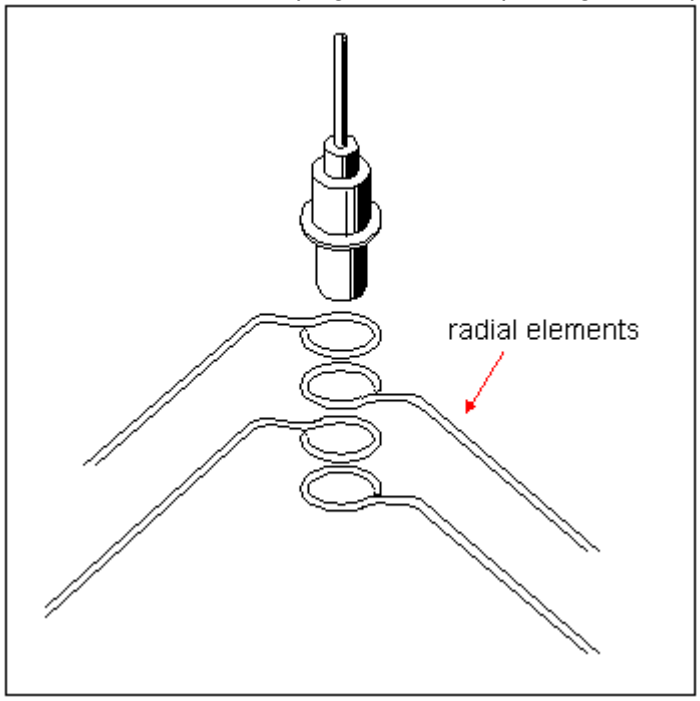


Figure 2.13 Assembling the radial elements into the antenna base portion.

Insert the remaining portion of the straight connector into the mounting bracket you made earlier sandwiching the radial elements between the bracket and the plain washer (see Figure 2.14). Secure the whole assembly by connecting the other PL-259 connector into the protruding part of the straight connector.

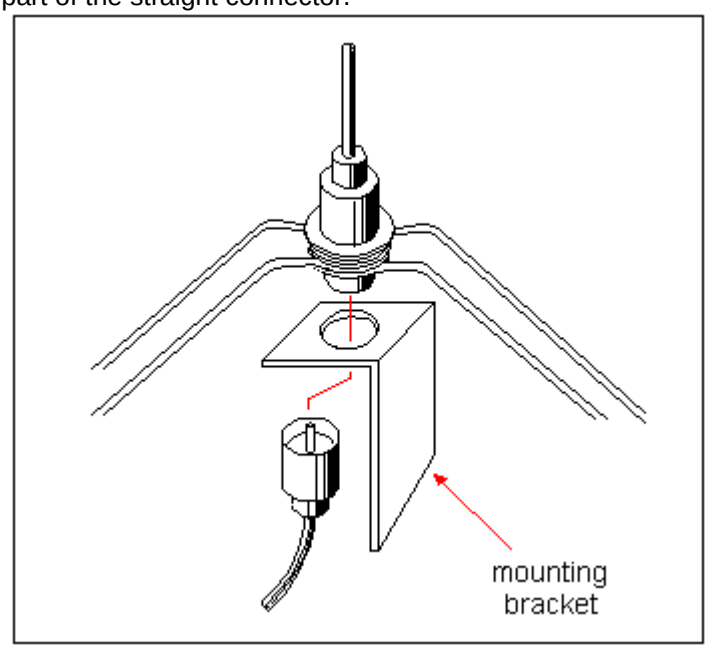


Figure 2.14 Assembling the antenna into the mounting bracket.

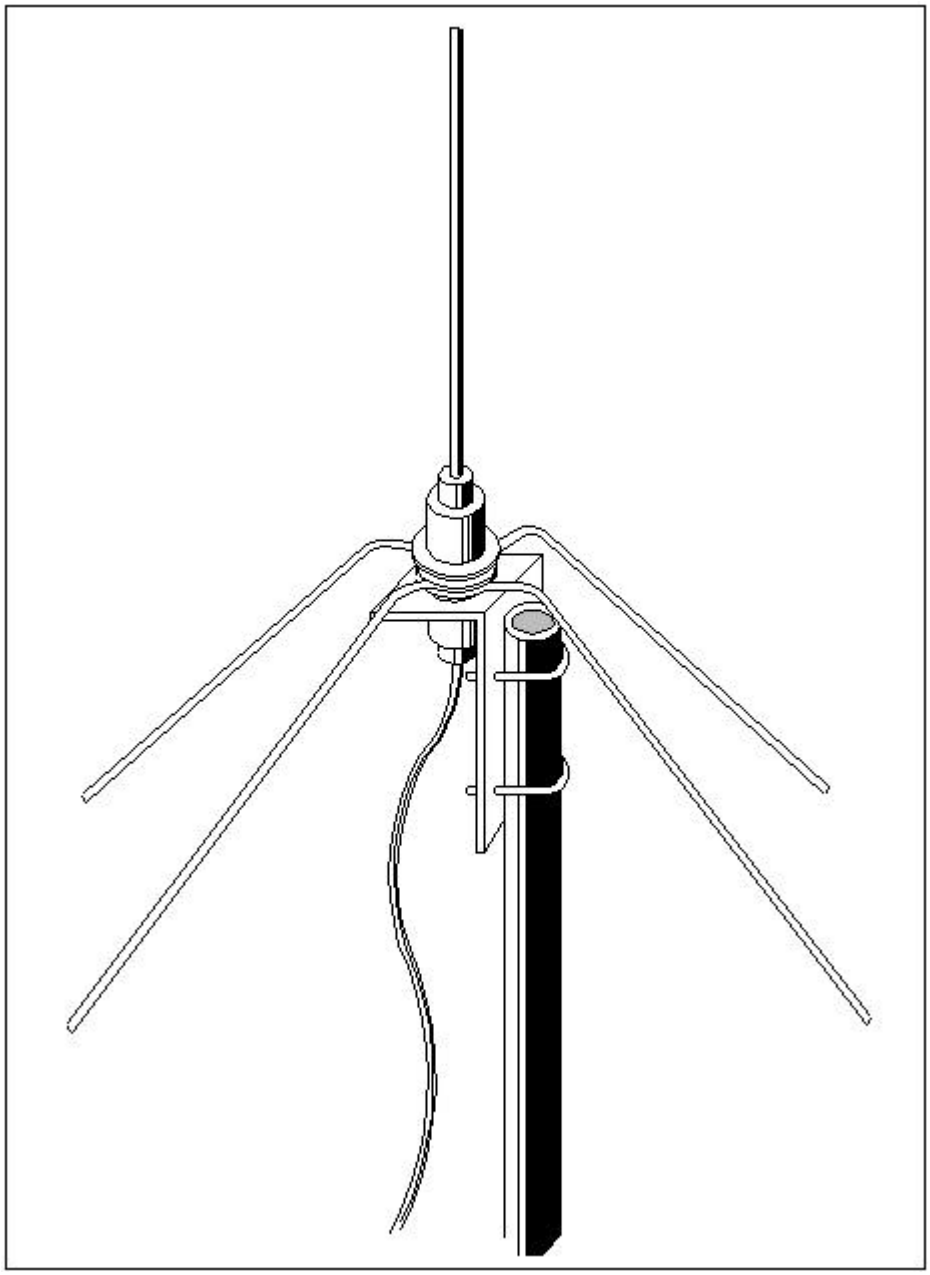


Figure 2.16 Mounting the groundplane antenna Model FQ-2.

Spread the radial elements around with equal spaces between them and tighten the PL-259 to fix the assemby firmly (see Figure 2.17).

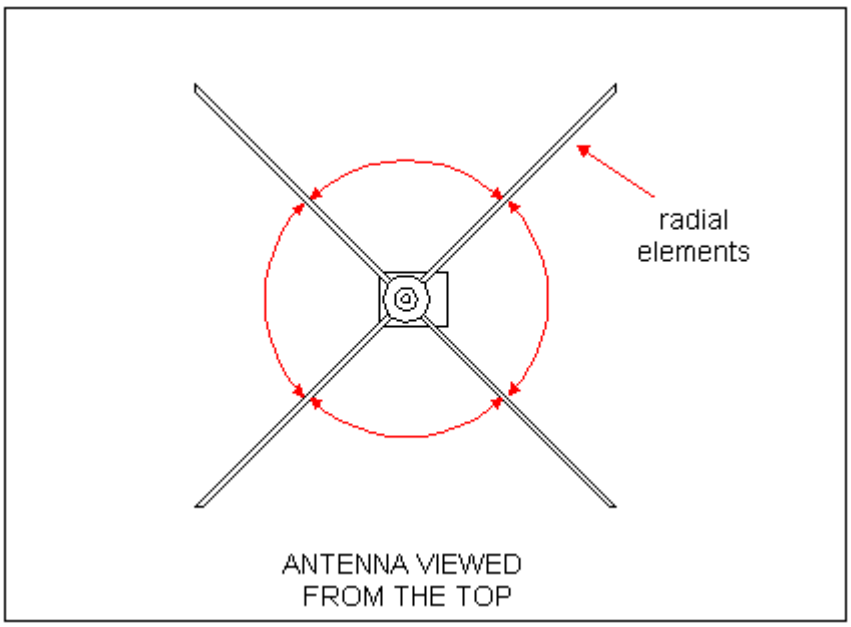
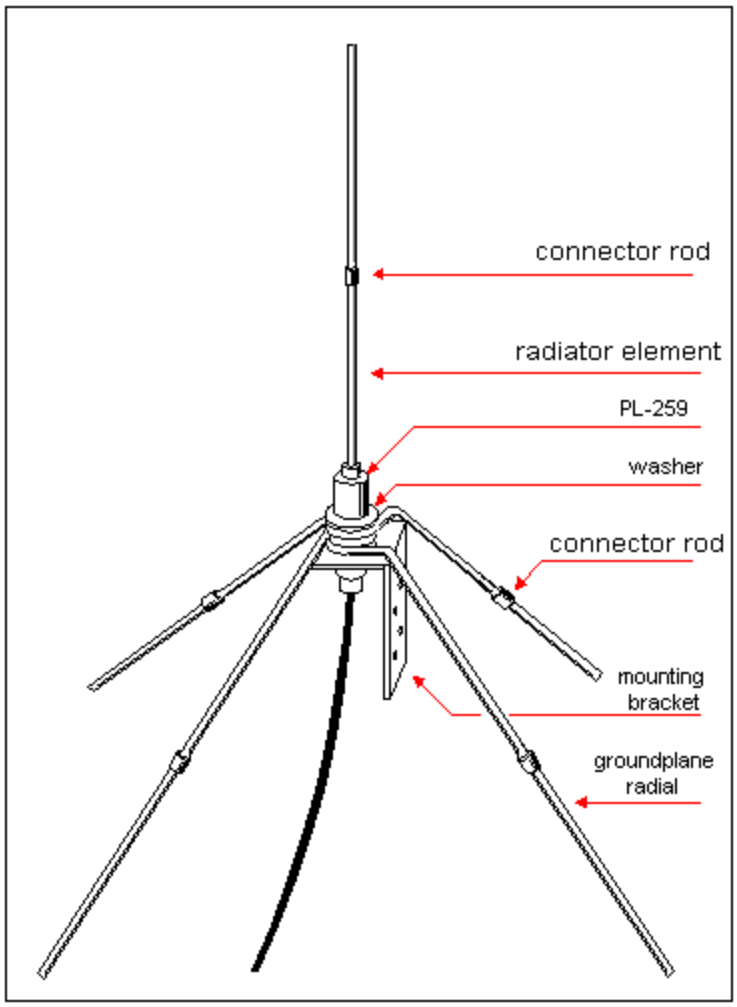


Figure 2.17 Spreading the radials.

3 GROUNDPLANE ANTENNA

Model FC- 2



The antenna is a vital link in the chain of radiocommunications and numerous designs have come off the drafting boards in a never ending search for improved performance. Experience shows that one major factor influencing the overall design of antennas is the particularity of the situation where it will be used. For example, the situation around fixed installations allows the antenna to be constructed with durable and heavy materials to make it mechanically strong. High power gain can also be easily attained by stacking a number of identical antenna. However, in mobile operation the situation drastically changes and using antennas designed primarily for fixed installations becomes impractical. Mobile operation imposes limitations on the design of an antenna regarding its weight, size, ruggedness, easiness in assembly and disassembly, and power gain. The operator has to choose a type of antenna which is highly portable and at the same time functionally efficient in mobile operations.

The groundplane antenna described in this chapter is another development from the FQ-2 model. It is actually the same antenna just "compacted" further to make its total size smaller and more portable when disassembled. This antenna was designed by a mobile radio operator several months after constructing his first groundplane antenna similar to Model FQ-2. Perhaps being unhappy about the bronze rods protruding out of his small knapsack he cut each rod in half and devised an ingenious way of connecting the elements together during assembly. That is how the FC-2 antenna was evolved.

The electrical characteristics of this antenna is similar to those of Model FQ-2. It also retains the mechanical durability of the earlier full length version. Being more compact it has become very popular among mobile radio operators.

Materials needed

The materials needed for Model FC-2 are the same with those needed for Model FQ-2. Refer to the preceeding chapter for the exact description of materials. The only difference between the two is the additional 3/16 inch diameter brass rod which is used to interconnect the detachable elements of the Model FC-2.

The compact detachable elements of the Model FC-2 permit it to be carried inside a pack or bag for mobile operations.

Construction

Fabricate a complete Model FQ-2 antenna following the construction methods described in the preceeding chapter. After you have constructed the Model FQ-2 disassemble it and cut the radiator and each radial element into two equal lengths (see Figure 3.2).

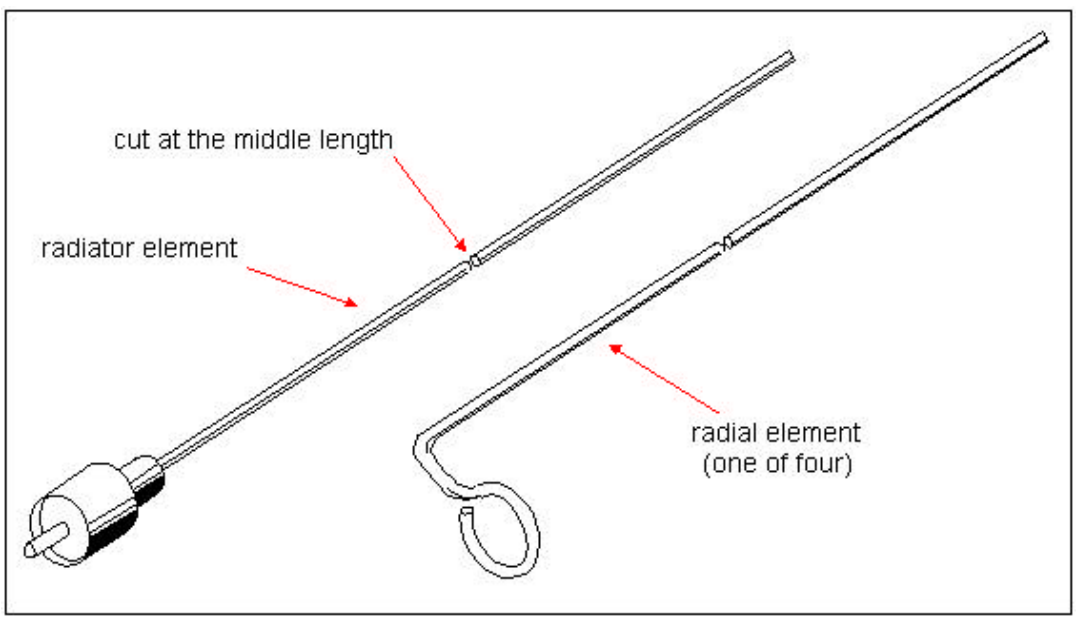


Figure 3.2 Cutting the elements into two equal lengths.

Next, take the 3/16" diameter brass rod and cut five 3/4 inch pieces from it. These short pieces of brass rod will be used to connect the two equal lengths of each element (see Figure 3.3).

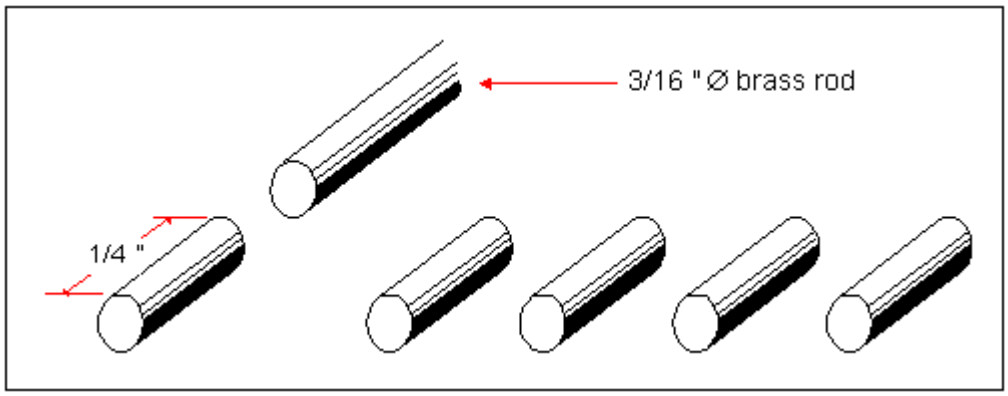


Figure 3.3 Preparing the connecting rods.

Drill a hole about 1/8 inch diameter at one end of each connector rod. The hole must be about half the length of the connector rod deep (see Figure 3.4).

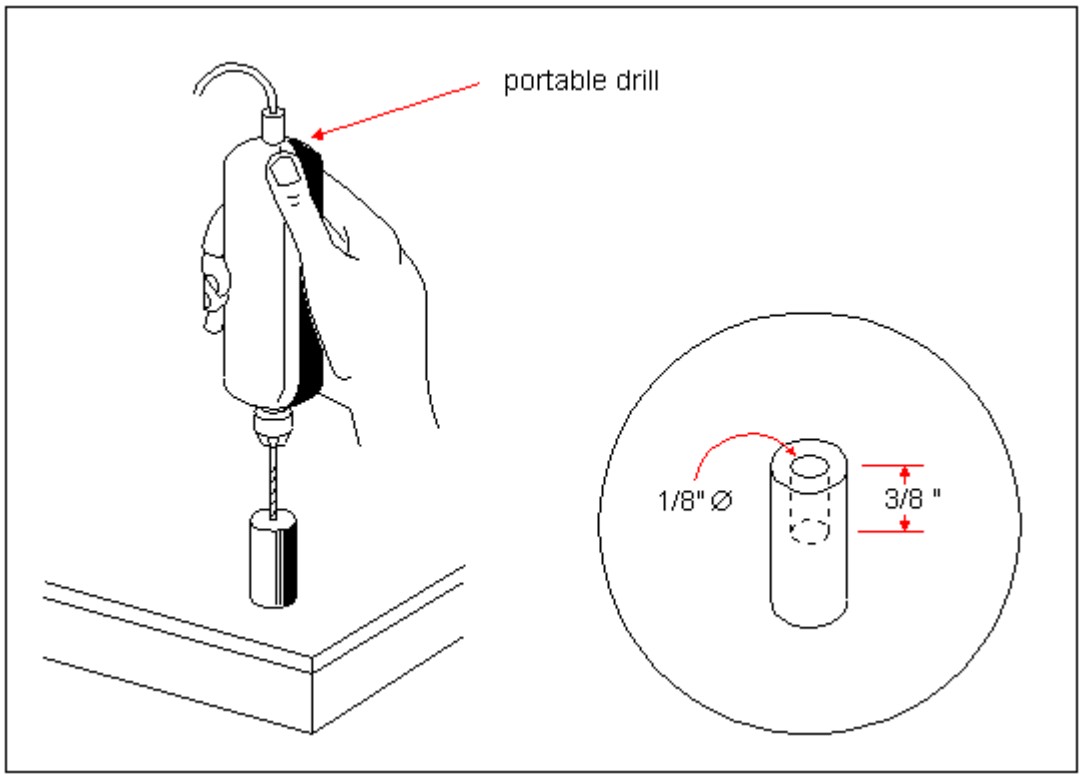


Figure 3.4 Drilling holes in the connector rods.

Next, drill another hole about 3/32 inch diameter at the other end of each connector rod - the same deepnes with the first hole (see Figure 3.5). Repeat the procedure for all five connector rods.

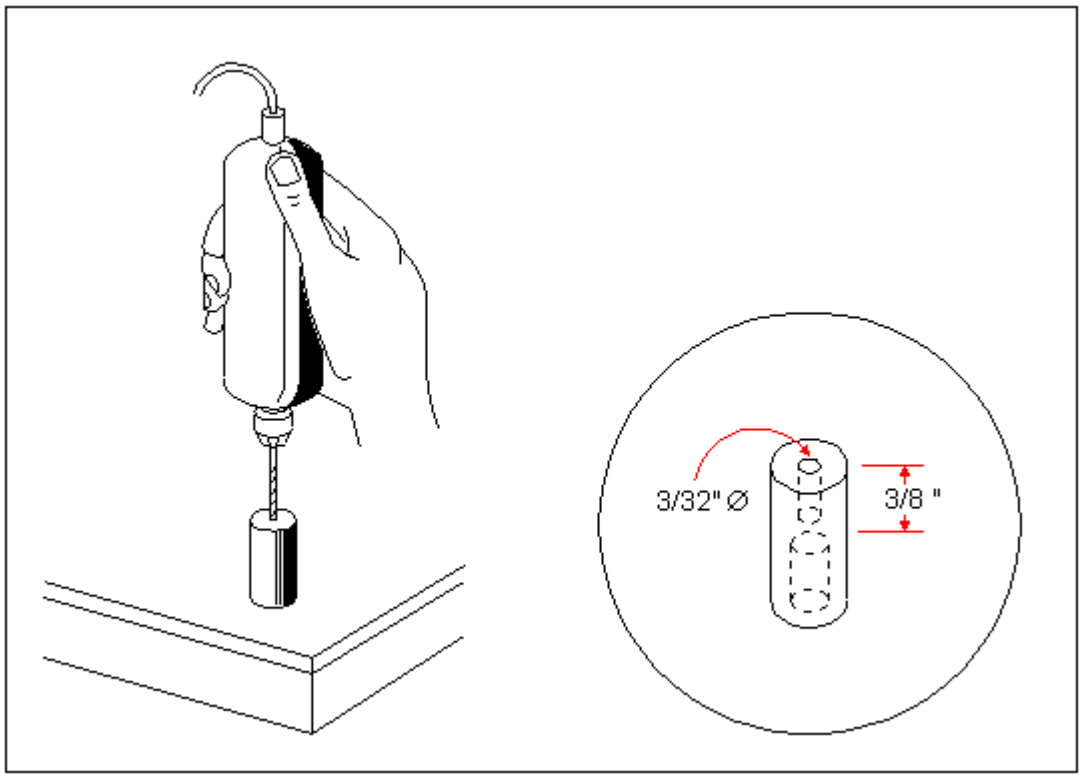


Figure 3.5 Drilling a 3/32" hole in the other end of the connector rods.

Next, secure the connector rod in a table vise and make a thread **inside the smaller hole** (the 3/32 inch hole to be sure) with a 1/8" gauge NF hand tap (see Figure 3.6). Repeat the same procedure for the remaining connector rods.

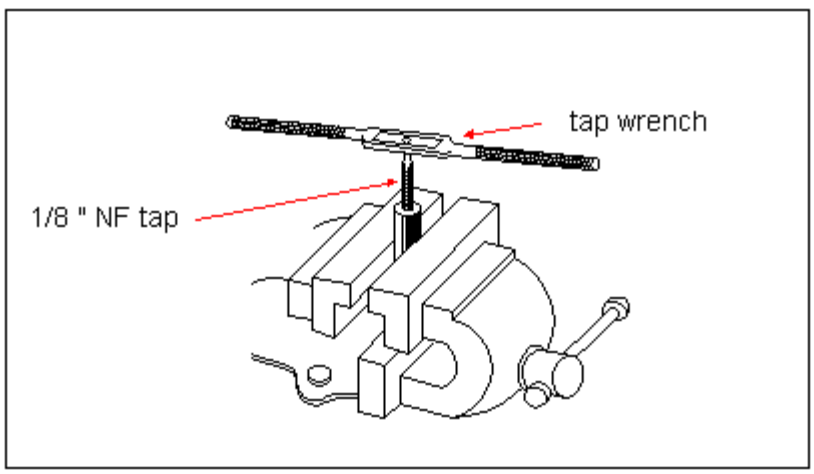


Figure 3.6 Making a thread in the smaller hole.

Insert the inner half rod of the radial element (the half part with the eyehook end) into the larger hole of the connector rod (1/8 " diameter hole unthreaded) and solder the two parts together. Do the same with the other radial elements (see Figure 3.7).

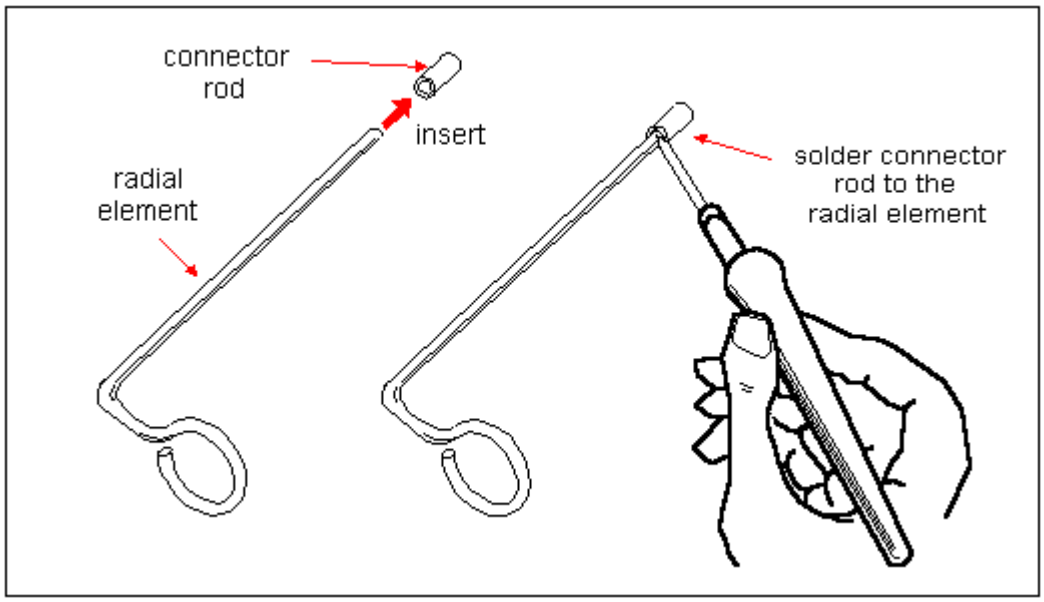


Figure 3.7 Coupling the connector rods to the radial elements.

Insert and solder the top half of the radiator element into the larger half of the remaining connector rod (see Figure 3.8). **NOTE: The purpose of this arrangement is to avoid the mistake of connecting the top half of the radiator to any of the radial elements.**

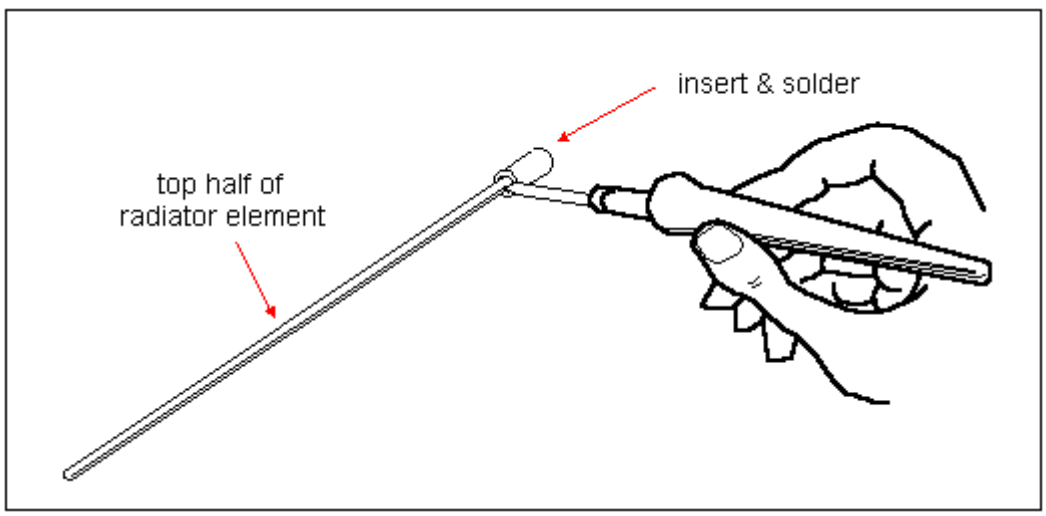


Figure 3.8 Coupling the top half of the radiator element to a connector rod.

Next step is to make a thread around one end of the outer half of the radial element. Use a manual threading die to make the thread. Secure the rod firmly in a table vise while threading. The thread must be at least 3/8 inch long. See Figure 3.9.

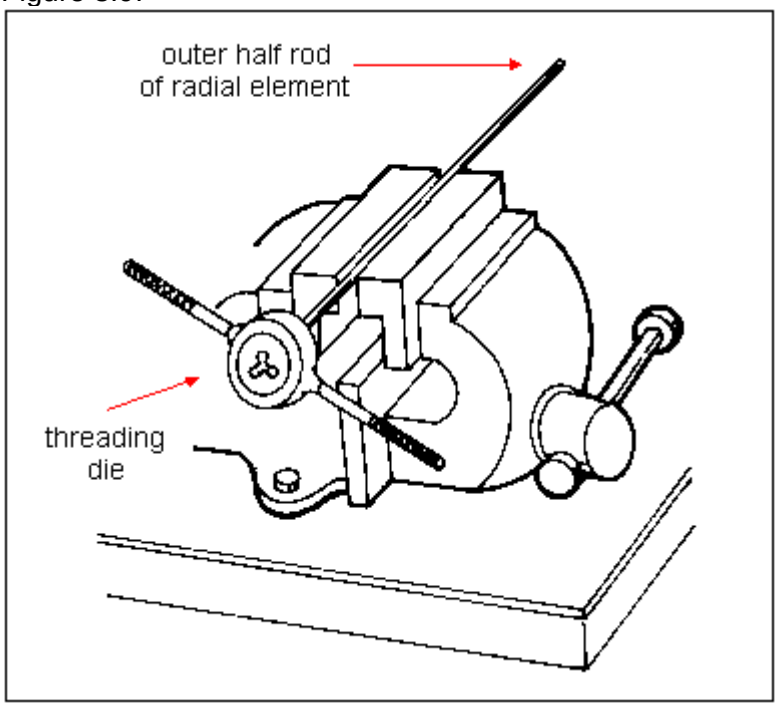


Figure 3.9 Making a thread at one end of the outer radial element.

After you have succesfully made the threads, screw each outer half into its respective connector rod (see Figure 3.10).

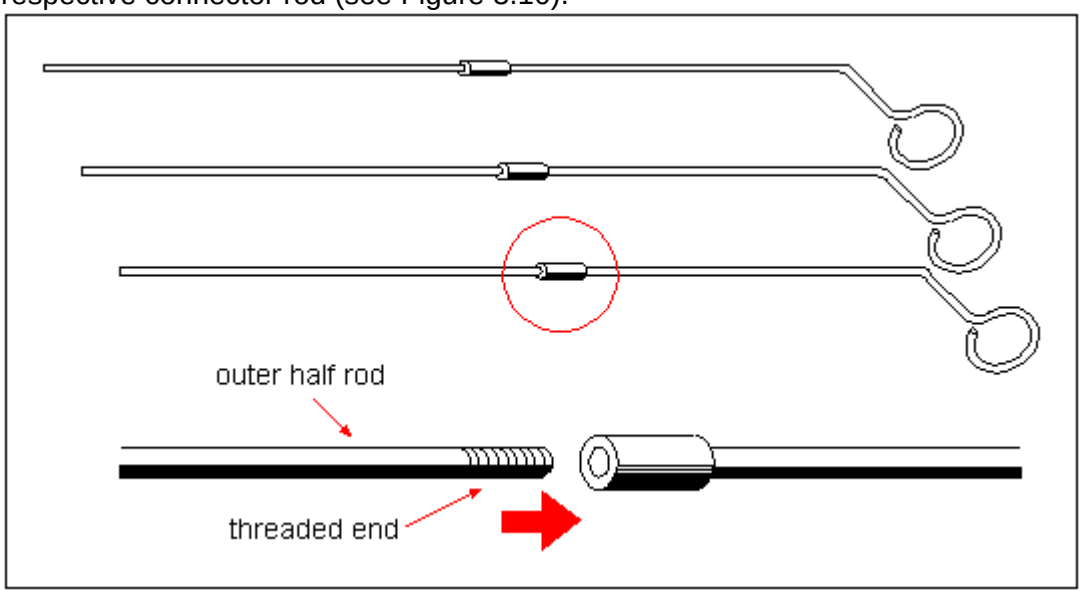


Figure 3.10 Assembling the radial elements.

Next, make a thread at the end of the lower half of the radiator element similar to what you have done to the radial elements (see Figure 3.11). Join the two halves of the radiator element together.

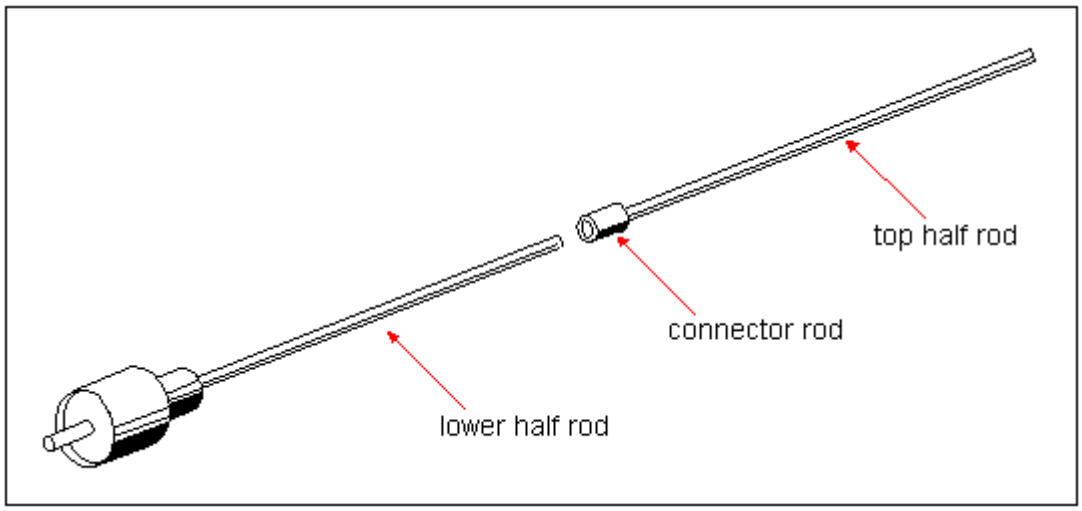


Figure 3.11 Assembling the radiator element together.

The final assembly of the Model FC-2 is similar to the Model FQ-2.

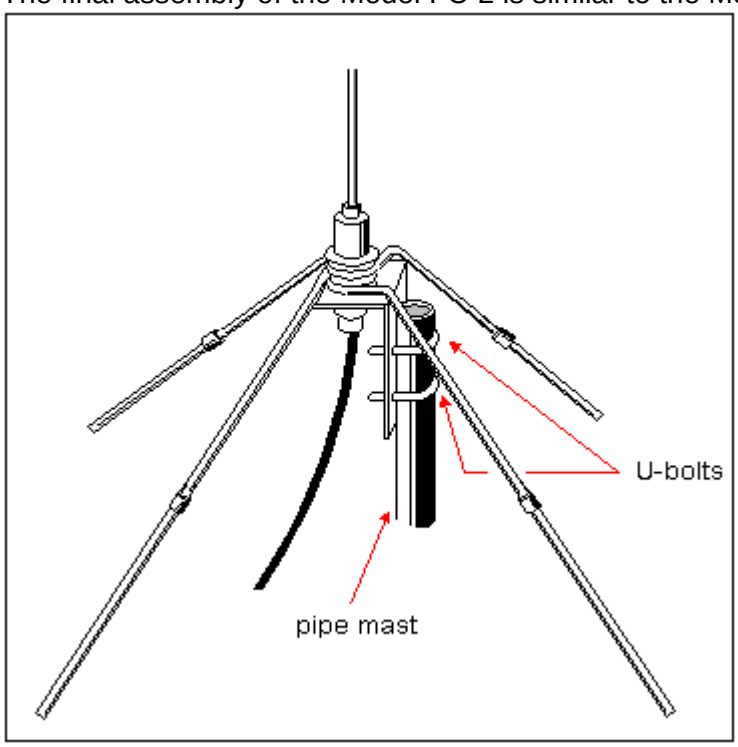


Figure 3.12 Final assembly and mounting of FC-2.

Mobile Installation

In mobile installations, the aluminum mounting bracket is not necessary and may be discarded and substituted with a 5/8" id* plain washer to hold the radial elements assembly. The antenna is then mounted by tying a rope at its base and hanging it under a tree or a makeshift post (see Figure 3.13a). An alternative method of hanging the FC-2 is to bend tip of the radiator element into a small hookform and a nylon rope is then tied to this hook to hang the antenna (See Figure 3.13b).

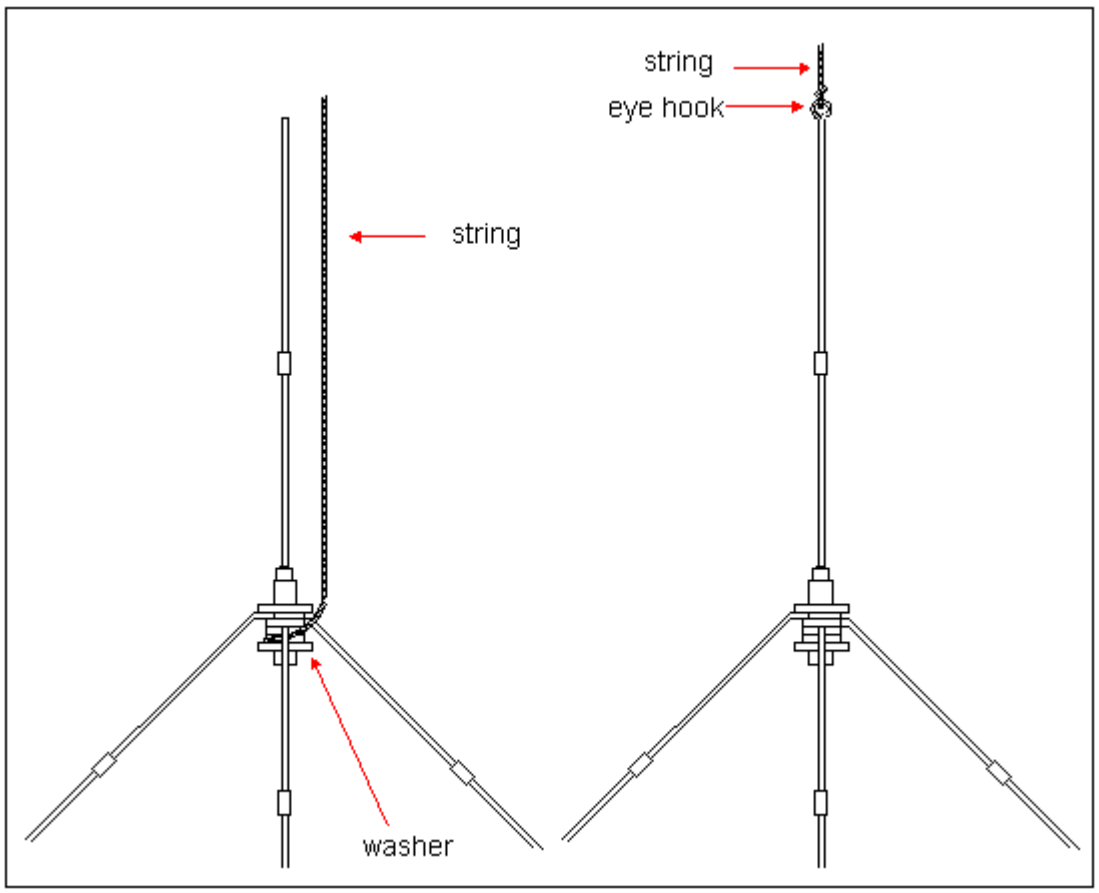
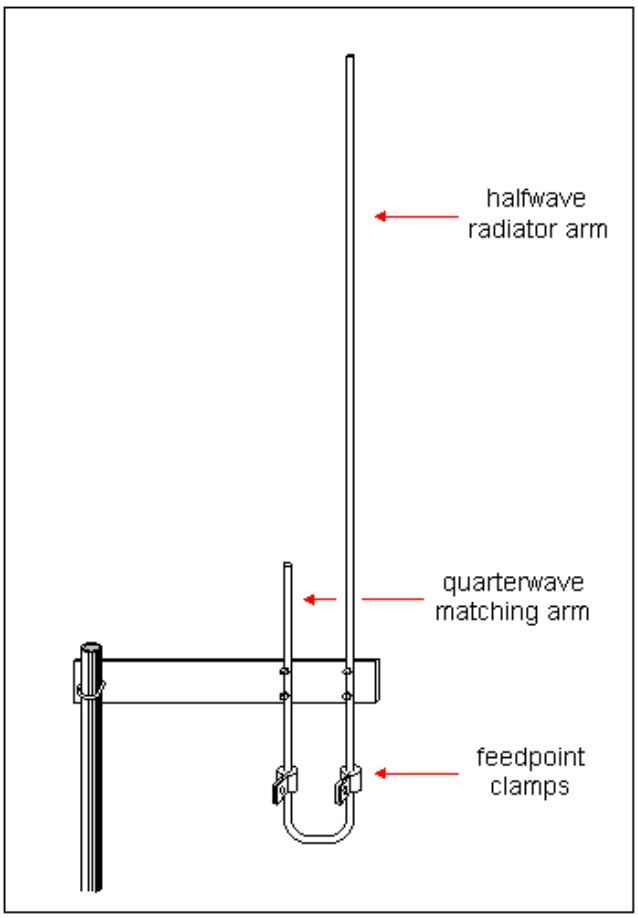


Figure 3.13 Mobile operation installation techniques.

4 J-FED HALFWAVE ANTENNA

Model JF-2



This antenna is specifically designed to satisfy the need for a simple but effective vertical antenna which does not require any grounding system. It is one version of a monopole antenna that carries its 'ground' along with it. The unit is comprised of a halfwave radiating element and a quarterwavelength matching section. The combination of these two elements provides the transformer action that matches the impedance. Although it is actually a quarterwave antenna, its radiation pattern and characteristics are very similar to those of a halfwave vertical antenna. It also exhibits a slight gain compared to a quarterwave groundplane antenna.

This antenna radiates its signal in an omni-directional pattern like most vertical antennas do. Its operational bandwidth is 140 - 150 MHz and exhibits an SWR response of less than 1.5:1 over the entire band.

The unit described in this chapter is designed for fixed installation. If you intend to use it for mobile operation it would be better if you modify the design to adapt it to the rugged environment it will encounter. Aluminum tubes in general are thin and soft and will easily crack if handled roughly so you must substitute it with brass, bronze or copper tubing. These materials are more expensive but they are more durable. They are also resistant to corrosion.

The elements must be cut in two or three sections and some means must be provided to join the pieces of tubing together in assembly (similar to FC-2 technique). You must also devise a method of mounting the antenna in a much simpler fashion than the one described here. Hanging the antenna under a tree or post will do, but there might be some other way that you can think of. One word of caution though, never use any metallic material to mount the antenna. All points in the antenna element are electrically active so it must be insulated from ground.

Materials List

Quantity	Specification/Description	Dimensions
1	Aluminum or Brass tube 3/8" od*	1" long
2	Aluminum strips - see text to make a strip out of a short length of aluminum tube	1/2" x 1-1/2"
1	Plastic plate 1/2" thick see text for details	3" x 12"
1	U-bolt with accompanying hex nuts and lockwashers	
4	Stove bolts - brass or GI with accompanying hex nuts and lock-washers	1/8" x 1"
2	Stove bolts - brass or GI	1/8" x 3/8"
2	Eye terminals - vinyl insulated	
4	Plain washers - 1/8" id**	
1	Hose clamp - enough to hold 1" diameter tube	

Miscellaneous: Epoxy glue

*od- outside diameter ** id- inside diameter

Construction

Cut the tube to a length of 81 inches using a suitable tube cutter. Next, starting from one end measure about 55 inches and starting at this point bend the tube to a U-shape. The two 'arms' of the bent tube must be spaced 2-1/8" apart from each other (see Figure 4.2).

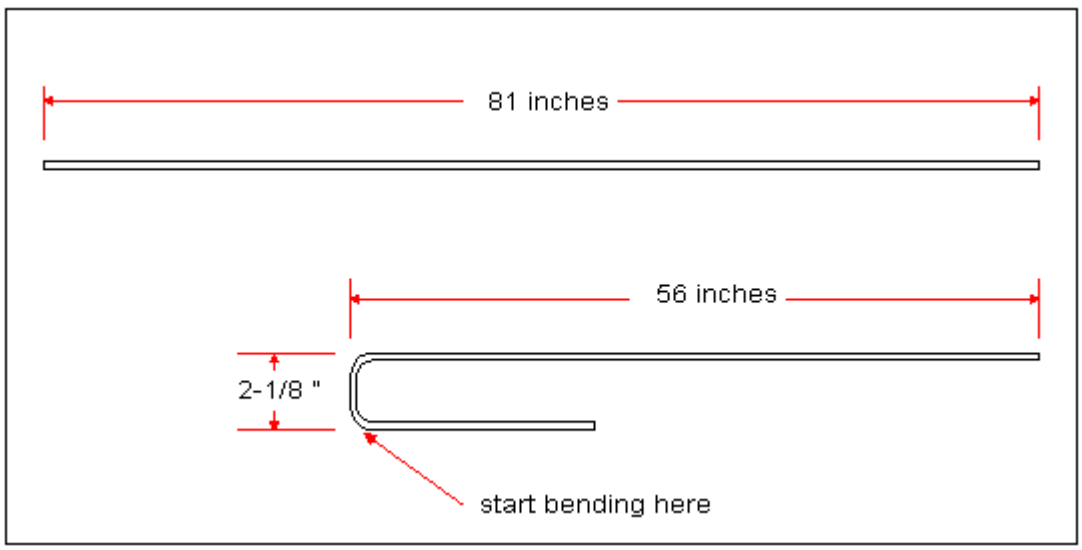


Figure 4.2 Bending one end of the tube.

Trim each arm of the tube to their proper lengths measuring from the extreme edge of the bend (see Figure 4.3). This method is employed to give an allowance for possible errors in bending the tube.

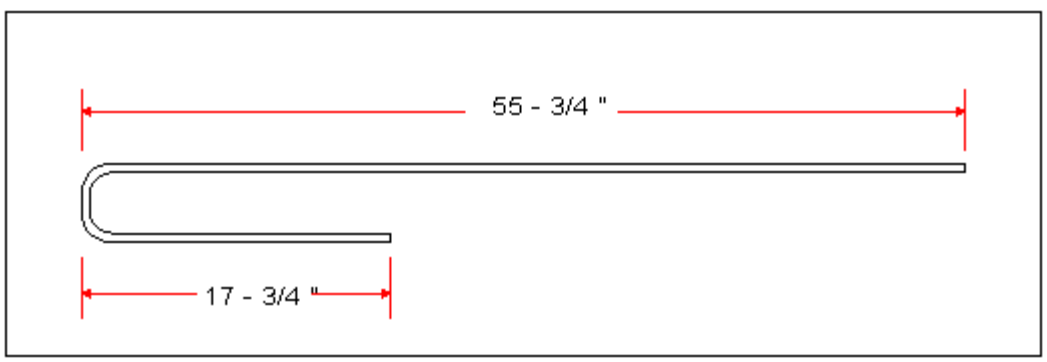


Figure 4.3 Trimming the tube to its exact length.

Drill four holes near the bend of the tube (see Figure 4.4). Each hole must be 1/8" in diameter.

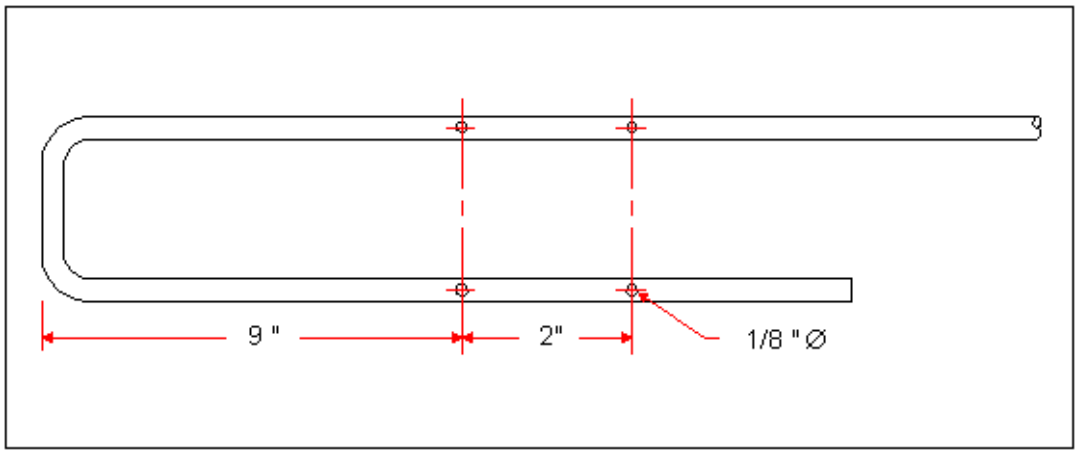


Figure 4.4 Drilling holes in the tube.

After drilling the holes, seal off both ends of the tube with an epoxy glue to avoid the entry of rainwater inside (see Figure 4.5). First, insert a substantial volume of cotton inside to act as a stopper for the epoxy. Then follow it up with epoxy glue levelling it to the edge of the tube. Let the epoxy set and dry before proceeding.

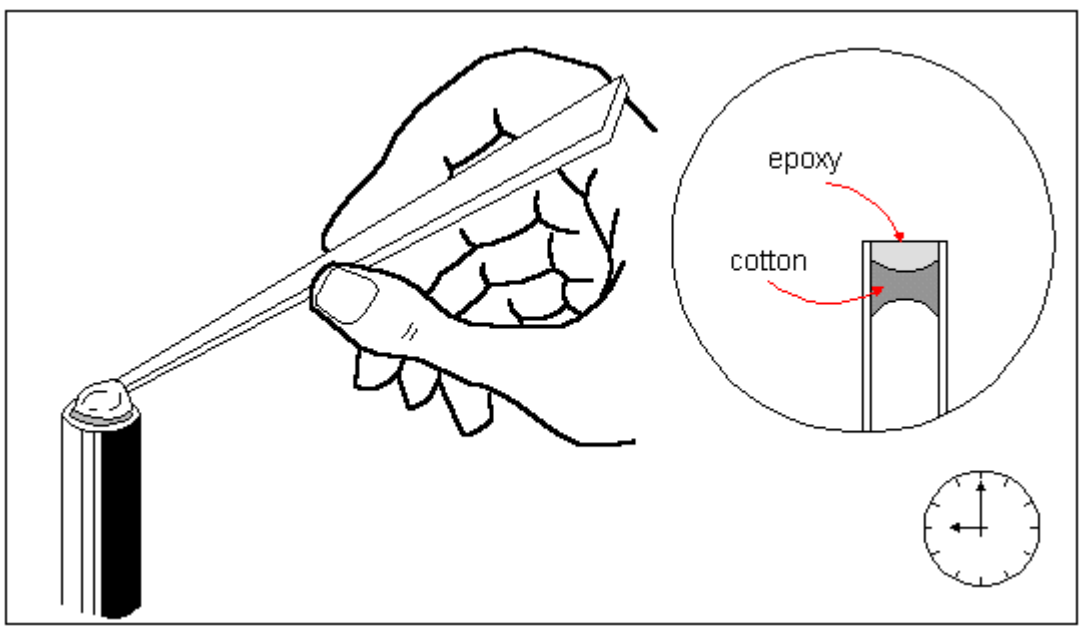


Figure 4.5 Sealing off the open ends of the tube with epoxy glue.

While you are waiting for the epoxy glue to dry, prepare the plastic plate for the antenna mount. Drill holes in the plastic plate following the dimensions shown in Figure 4.6. The larger hole (3/16 " diameter) is intended for the U-bolts so their dimensions must coincide with the actual U-bolt used.

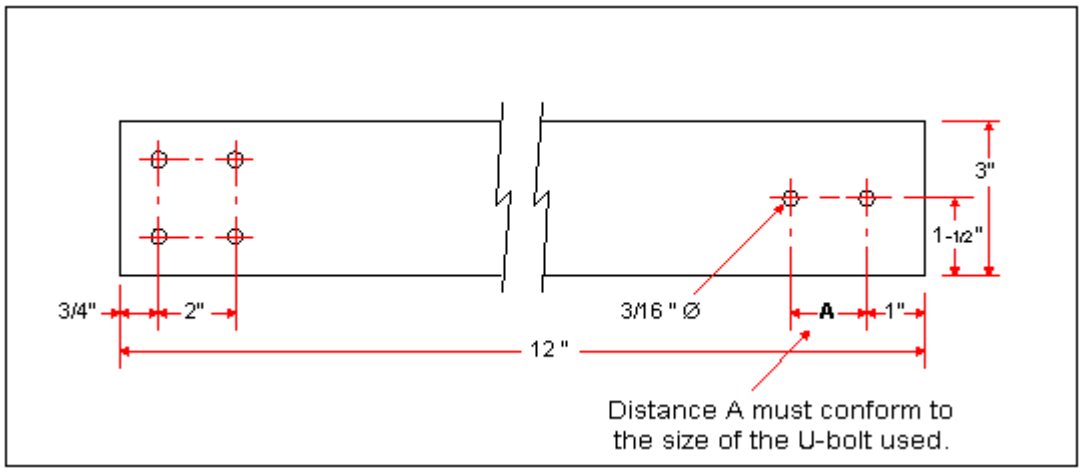


Figure 4.6 Preparing the plastic mounting plate.

Fabricate a metal strip out of a short length of aluminum tube (about 5 inches long) by pressing it in a table vise until the tube is flattened (see Figure 4.7).

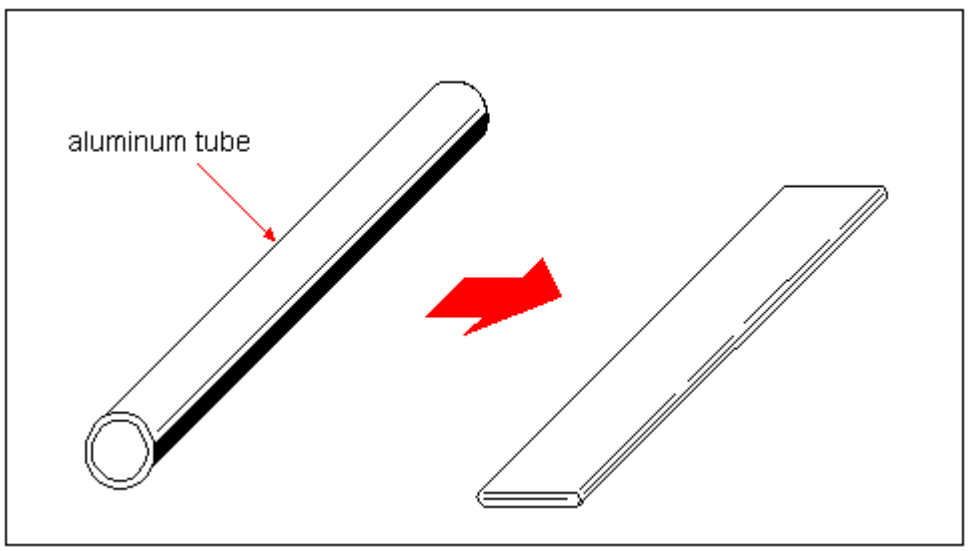


Figure 4.7 Fabricating a metal strip out of a short aluminum tube.

Out of this strip cut two short pieces (about 1-1/2" long). Bend the two strips to a form of a clamp to fit tightly around the antenna tubing (see Figure 4.8). These clamps serve as the feedpoint terminals of the antenna.

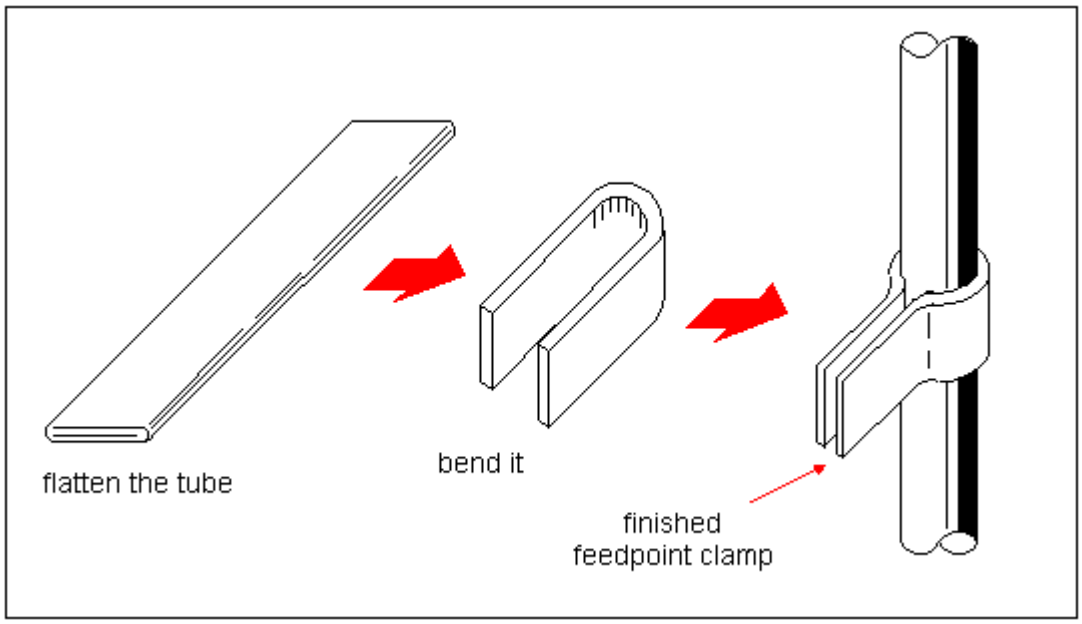


Figure 4.8 Fabricating the clamps.

Next, drill a hole about 1/8" diameter through the flattened end of each feed-point clamp (see Figure 4.9).

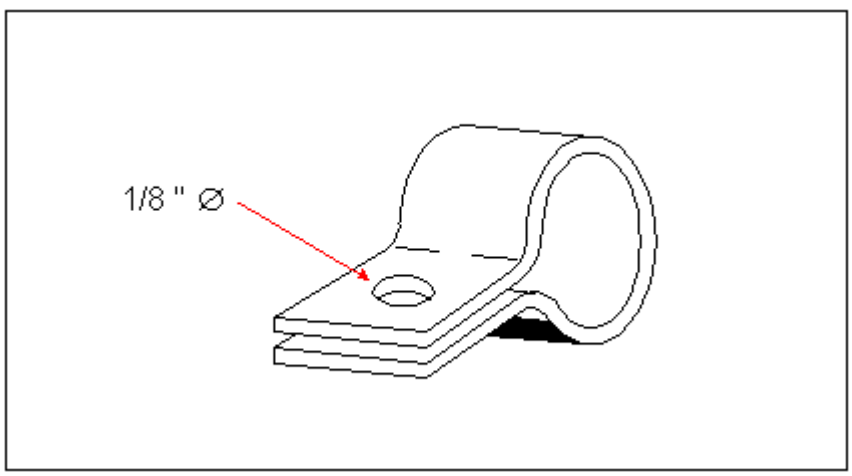


Figure 4.9 A feedpoint clamp with a drilled 1/8" diameter hole.

Assembly

First, attach the J-shaped tube to the plastic mounting plate with 1/8" x 3/4" stove bolts made of corrosion proof materials such as brass or stainless steel. Do not forget to include the necessary lockwashers in the attachment (see Figure 4.10). Be careful in tightening the nut because the tube is hollow inside and it might collapse damaging the tube. Apply torque to the nuts just enough to hold the tube rigidly.

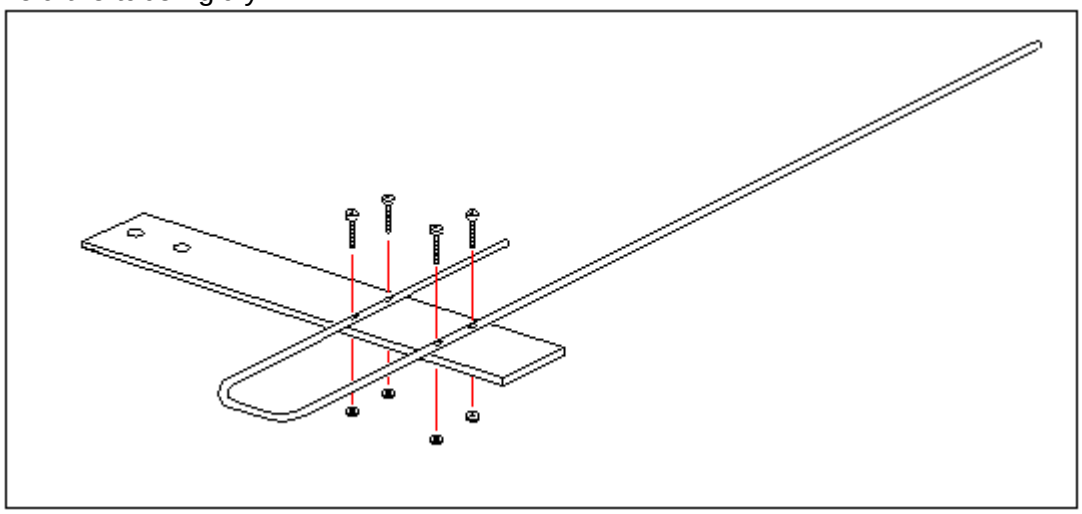


Figure 4.10 Securing the J-shaped tube to the mounting plate.

Attach the feedpoint clamps into both arms of the the tube. Attach one clamp on the shorter arm of the tube and attach the other clamp on the longer arm (see Figure 4.11).

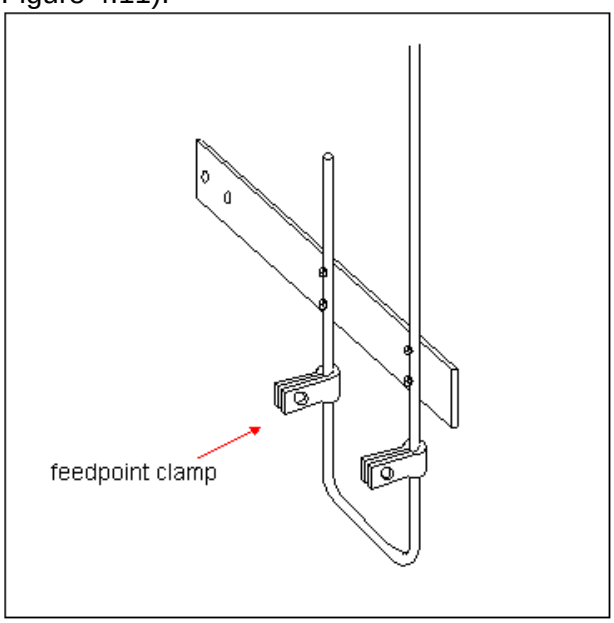


Figure 4.11 Feedpoint clamps attached to the antenna.

Attach a plain washer and an eye terminal into a stove bolt (1/8" x 3/16") then insert the bolt into the hole in the feedpoint clamp sandwiching the eye terminal inbetween (see Figure 4.12). Place a lockwasher and a hex nut at the other end of the bolt then tighten the clamp lightly. Repeat the same procedure for the other clamp.

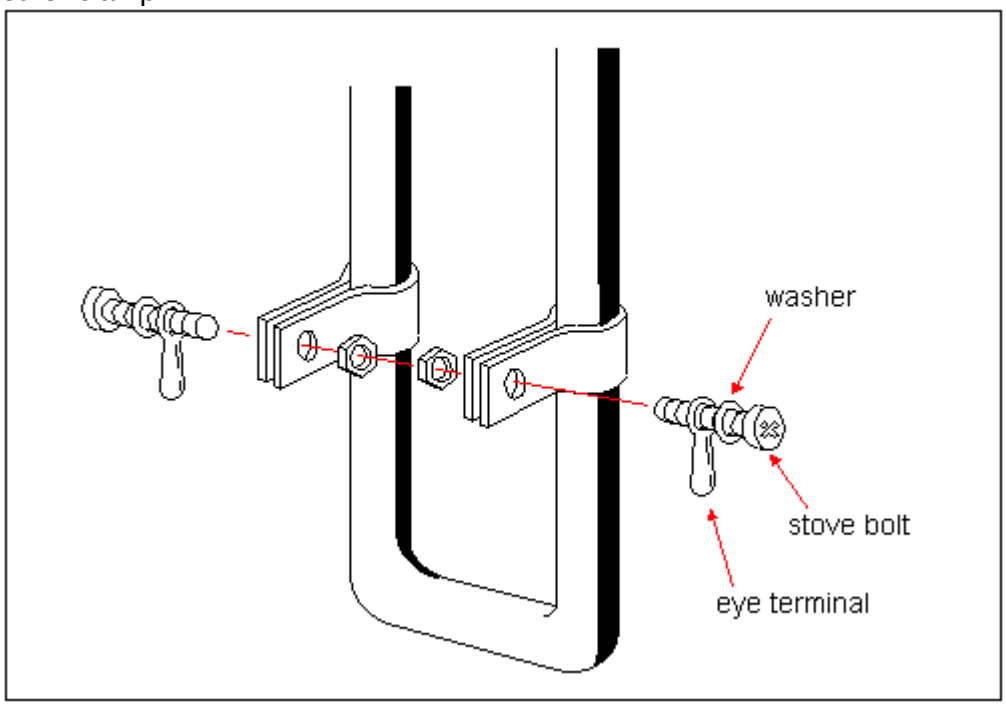


Figure 4.12 Assembling the feedpoint terminals.

Next step is to connect the coaxial cable to the feedpoint terminals. Prepare one end of the coaxial cable by separating the braid/shield from the inner conductor (see Figure 4.13).

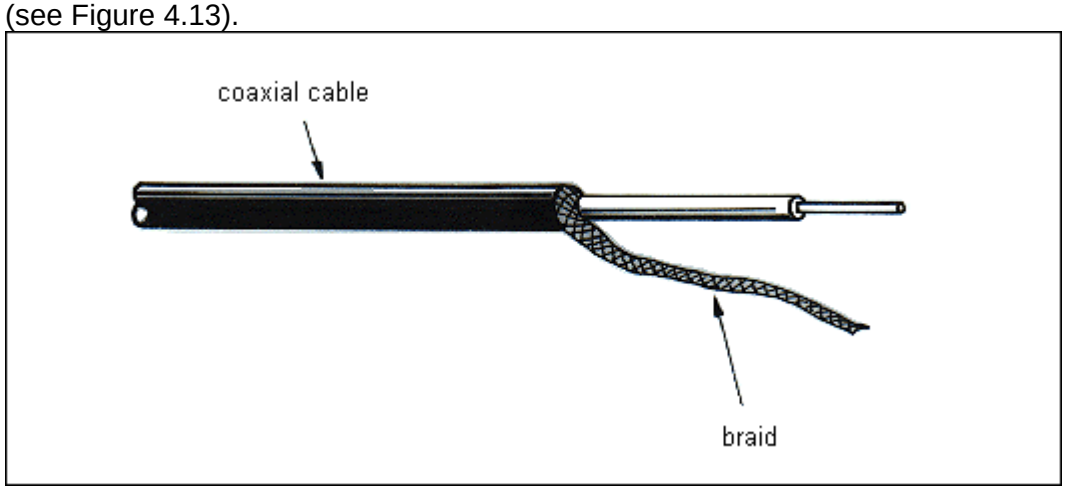


Figure 4.13 Making a pig tail.

Insert and solder the two conductors (braid and inner conductor) to the eye terminals attached in the feedpoint clamps. The braid must be connected to the shorter arm of the tube and the inner conductor must be connected to the longer arm (see Figure 4.14).

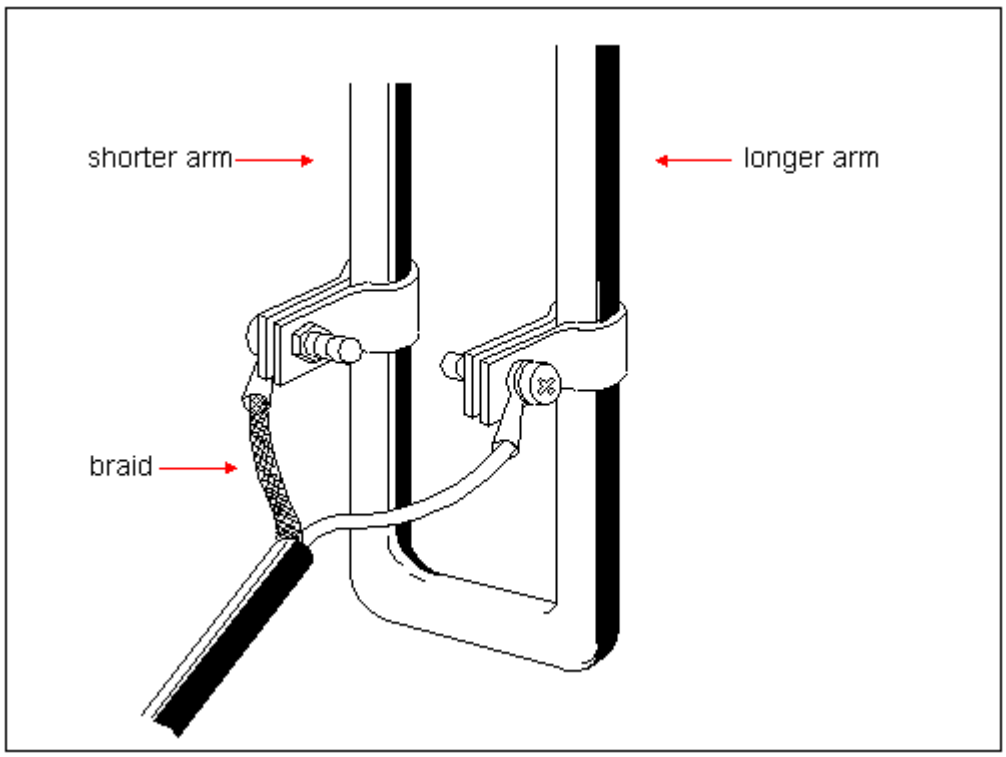


Figure 4.14 Connecting the coaxial cable to the feedpoint clamps.

Mount the antenna to the mast you intend to use. It is best to tune the antenna to resonance right at the mast where it will be installed permanently. Connect the coax cable to an SWR meter. The coaxial cable must be furnished with the right connectors for the particular type of SWR meter you use. Connect a transceiver to the input connector of the SWR meter (usually marked 'transmitter'). Set the transceiver to 145.00 MHz and key the PTT to transmit. Note the SWR reading on the meter. While the transceiver is on standby , move both feedpoint clamps higher or lower than the initial setting until you get a low SWR response over the entire frequency range (140.00 MHz to 150.00 MHz) specifically). Move the clamps about 1/4" at a time (see Figure 4.15).

IMPORTANT:

Do not move the clamps while the rig is transmitting and do not touch any part of the antenna when reading the SWR response. The position of the clamps must be moved always at the same level at the same time.

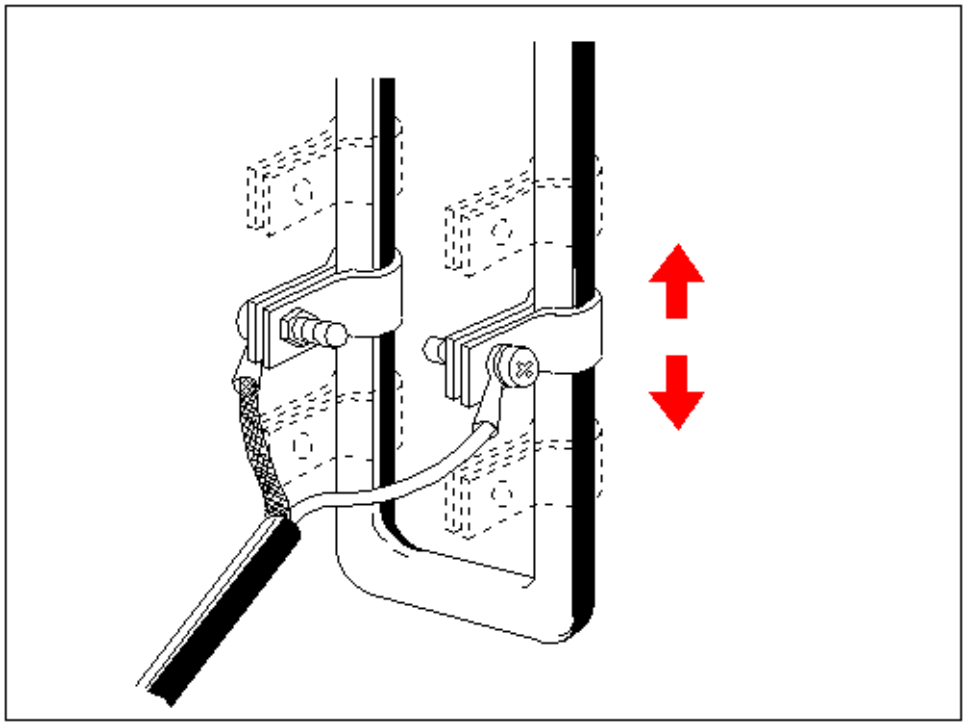


Figure 4.15 Adjusting the position of the clamps to tune the antenna.

After you have tuned the antenna to resonance, tighten the nuts holding the feedpoint clamps permanently and fix the coaxial cable to the mounting plate with plastic clamps (see Figure 4.16).

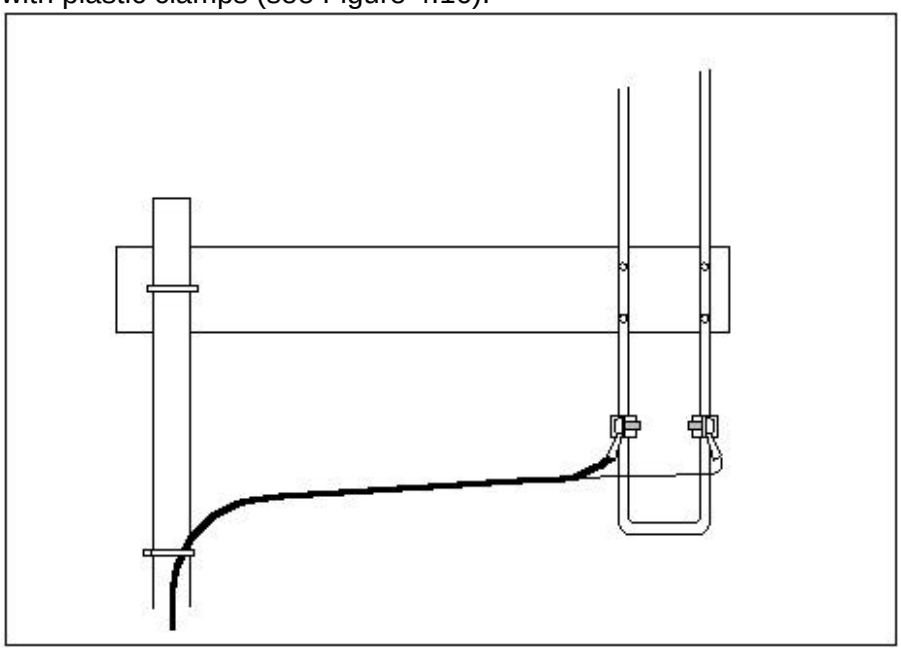


Figure 4.16 Final mounting of the J-fed antenna.

5 COAXIAL DIPOLE

Model CD-2

The coaxial dipole described here has the advantage of having lower resistance to wind compared to the groundplane designs. It has also a narrow form which some radio operators find beautiful. The following illustrations in Figure 5.1 show how the coaxial dipole was evolved from a basic dipole antenna.

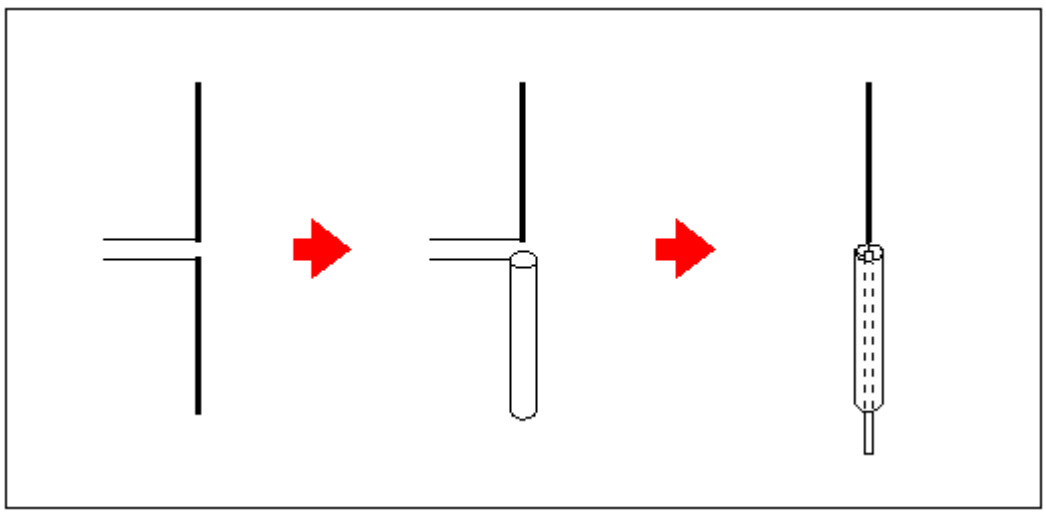


Figure 5.1 Evolution of the coaxial dipole from a basic dipole antenna.

As you can see in the illustration, one of the elements is enlarged to form a tube. The coaxial transmission cable is then inserted through this tube, with the inner conductor of the coaxial cable connected to the radiating element and the shield connected to the tube. The tube functions as a groundplane.

The CD-2 coaxial dipole has an operational bandwidth of 140-150 MHz. It exhibits an SWR response of less than 1.5:1 over the entire band. It has a power gain of 1 dB (unity gain) compared to a standard dipole reference. The RF signal radiates from the antenna in an omni-directional pattern. Likewise, it recieves signal equally well from all directions. This unit is designed to be installed primarily in base stations but it could be used for mobile applications too. The radiating element must be detached when transporting the antenna.

SCALED DOWN APPEARANCE

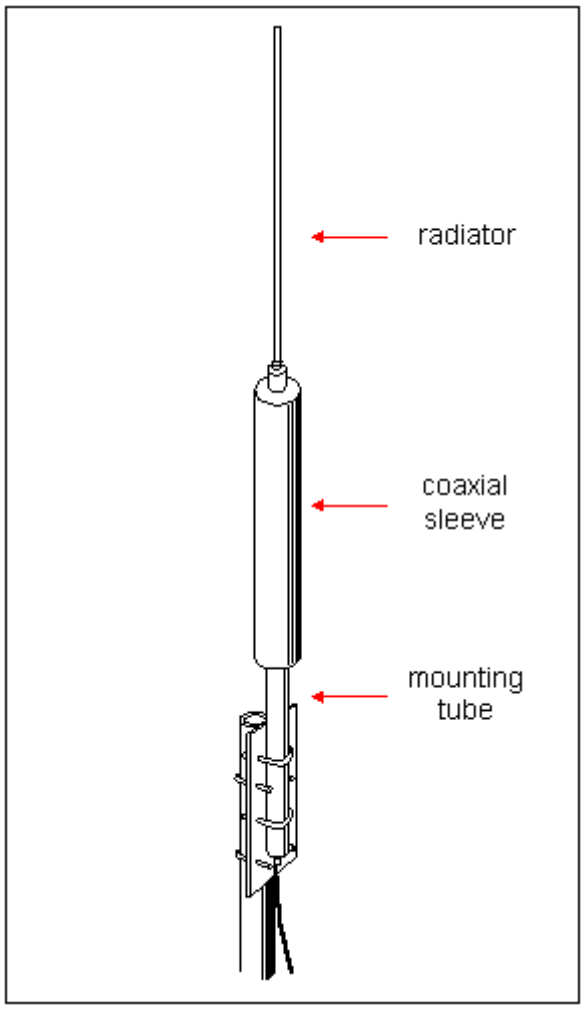


Figure 5.2

Materials List		
Quantity	**Specification/Description**	**Dimensions**
1	Aluminum tube	3" x 18"
1	Aluminum tube	1" x 36"
1	PL-259 VHF male connector	
2	SO-239 VHF female connector	
1	Brass rod 1/8" od* - the brass rod for acetylene welding is recommended	
2	Aluminum bushing - see main text for exact dimensions	
4	U-bolts - with accompanying hex nuts and lockwashers	
1	Aluminum Plate or GI	3" x 6"
9	Self tapping metal screws	1/8" x 1/2"
1	Short length of coaxial cable RG-58/U	37" long

*od- outside diameter

Construction

First prepare the two aluminum tubes of different diameters. Cut the tubes to their proper lengths as shown in Figure 5.3

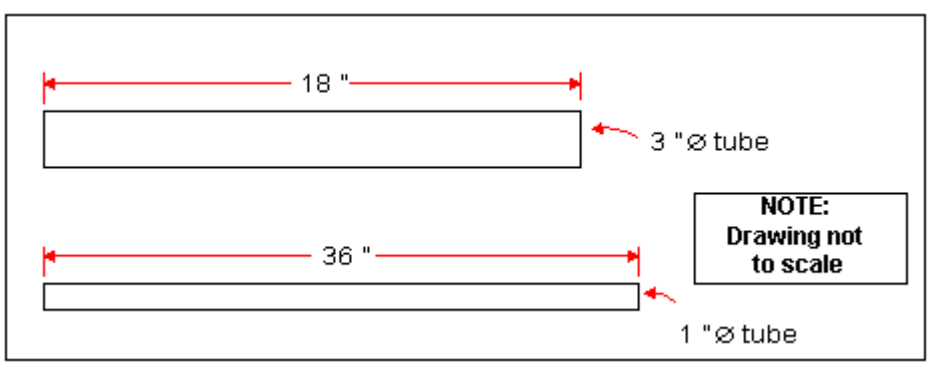


Figure 5.3 Cutting the tubes to their proper lengths.

Next, drill three holes (1/" diameter) at **both ends** of the longer tube. The holes must be equally spaced from each other (see Figure 5.4).

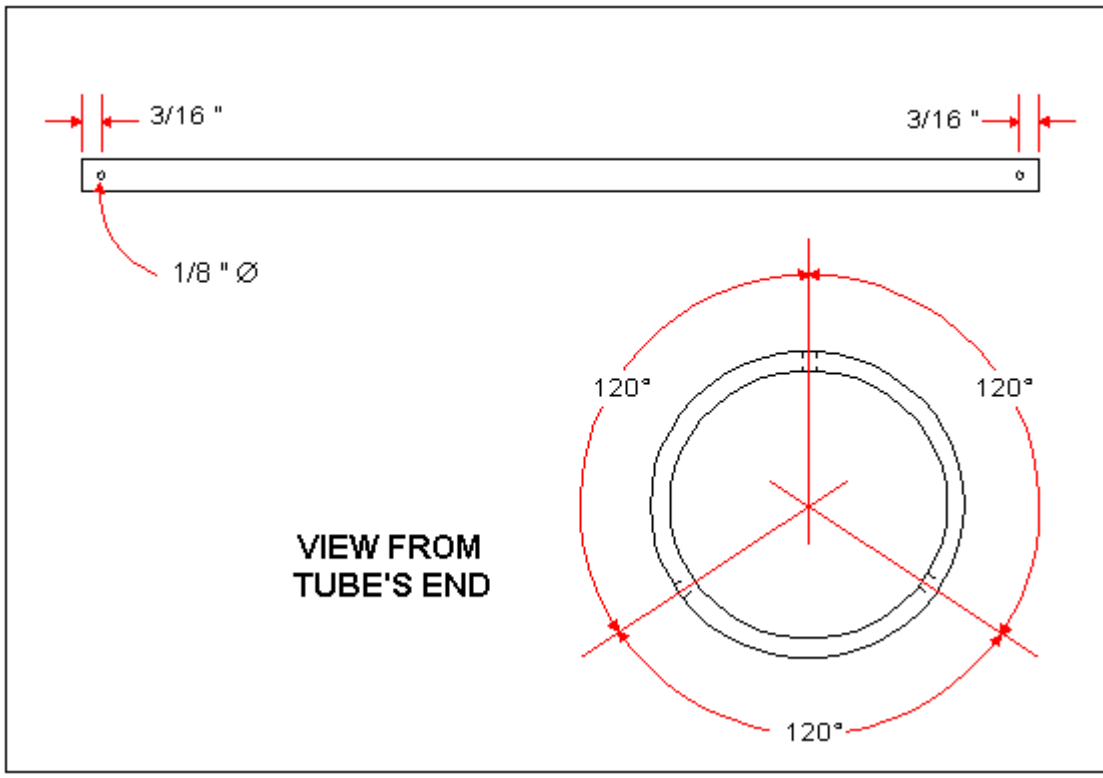


Figure 5.4 Drilling three holes at both ends of the long tube.

Drill three holes (1/8" diameter) at **one end** of the shorter tube. The holes must be 1/4" away from the edge and equally spaced from each other (Figure 5.5).

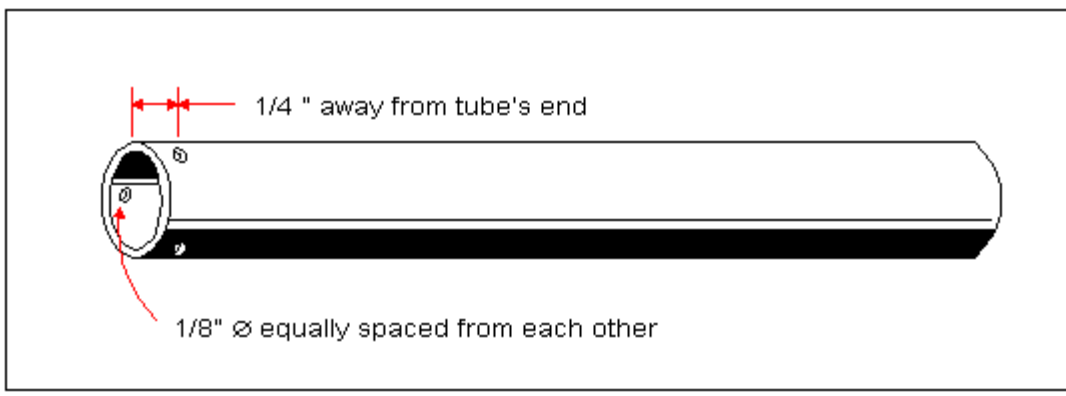


Figure 5.5 Drilling a hole at one end of the short tube.

Machine the smaller bushing from a thick aluminum slab or rod to its proper size. Follow the dimensions in Figure 5.6.

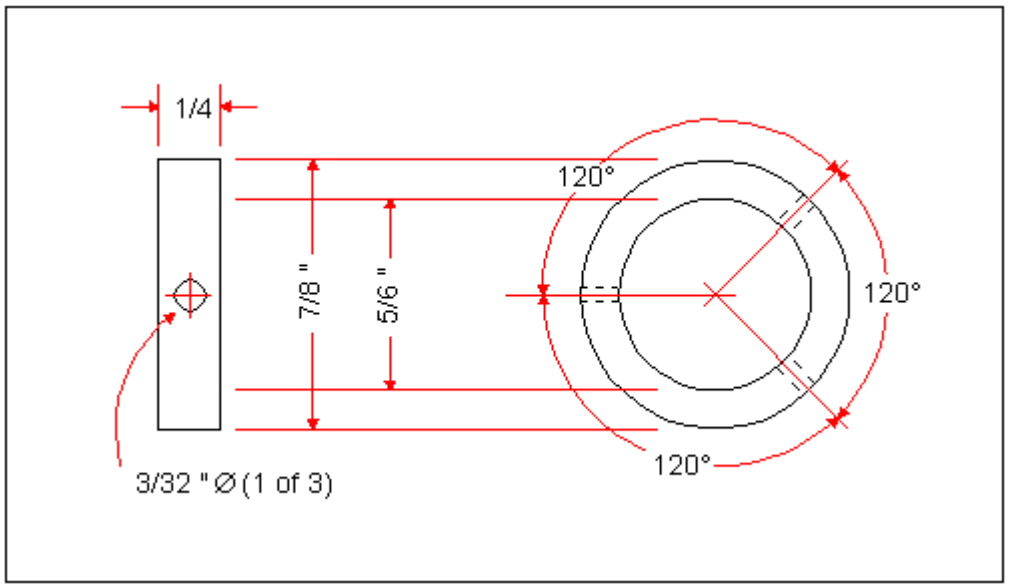


Figure 5.6 Smaller bushing dimensions.

Next, machine the larger bushing from similar material. Follow the dimensions shown in Figure 5.7.

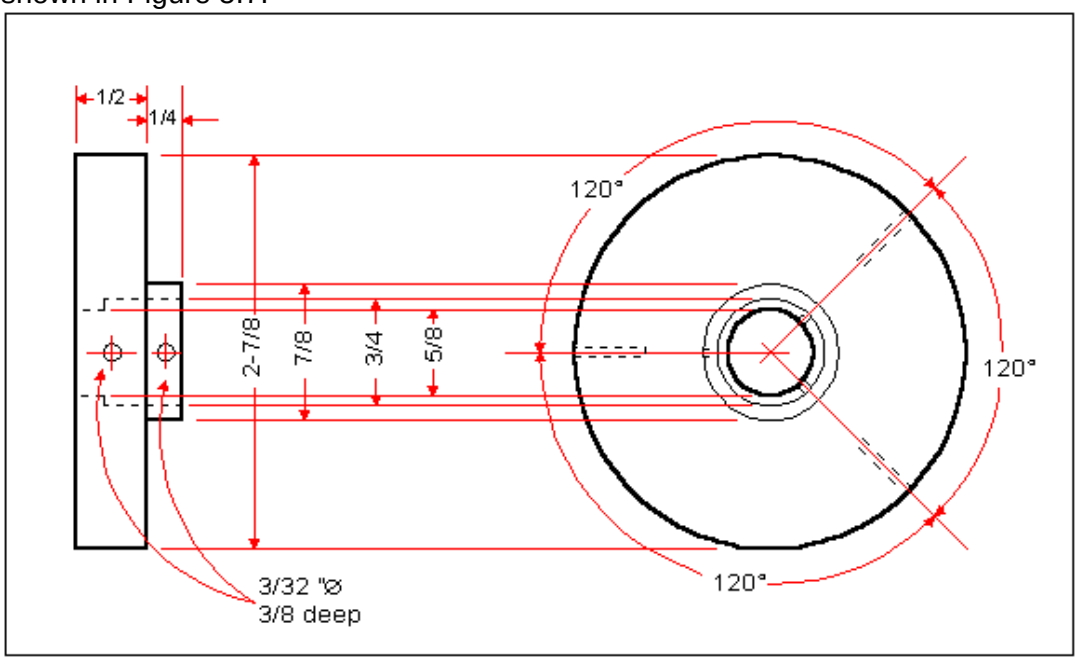


Figure 5.7 Dimensions of the larger bushing.

File away a small portion at one end of the brass rod reducing it to smaller diameter enough to fit inside the center pin of PL-259 connector. Solder the brass rod into the center pin of the PL-259 connector (see Figure 5.8).

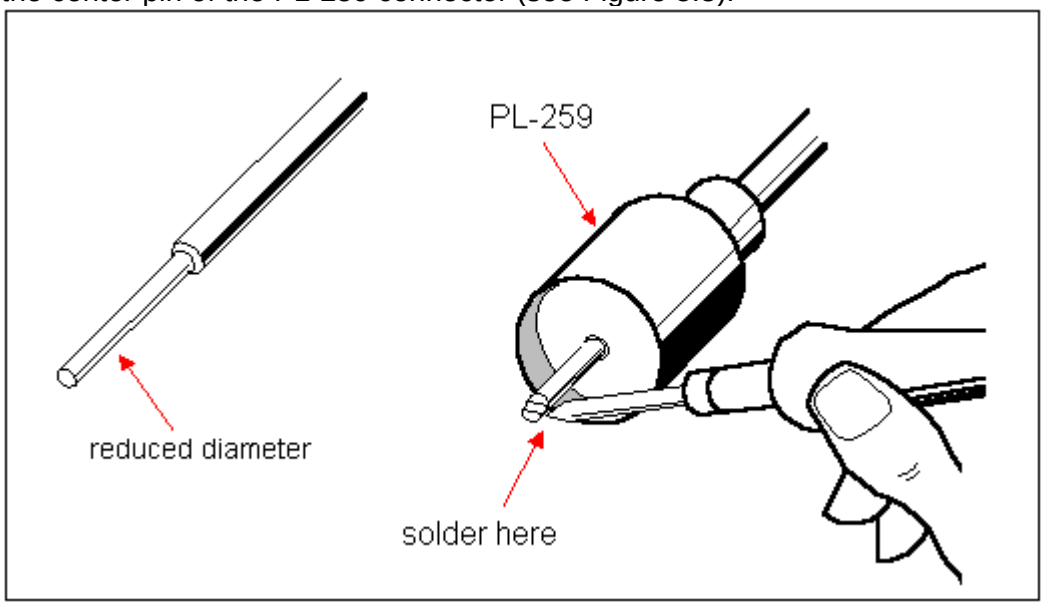


Figure 5.8 Soldering the radiator element to the PL-259.

Cut a small length of coaxial cable (about 2 inches) and remove its inner conductor and braid/shield. You need only the vinyl outer jacket. Insert it into the brass rod all the way inside the PL-259 connector. Cut away any protruding vinyl portion. The jacket serves as an insulator between the brass rod and the body of PL-259 (see Figure 5.9).

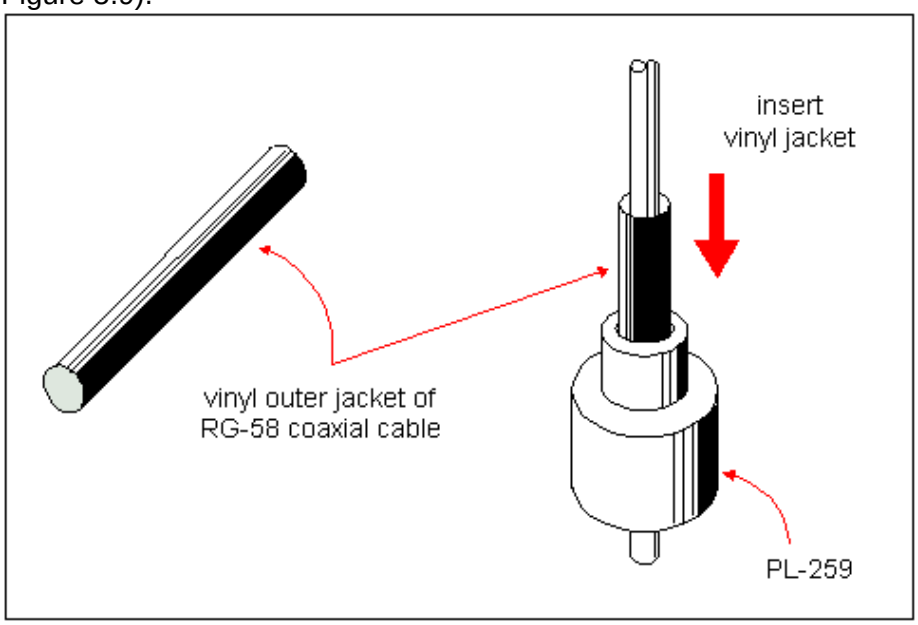


Figure 5.9 Inserting the insulating jacket into the PL-259.

Mix equal amount of epoxy glue and place it over and around the protruding part of the vinyl jacket (see Figure 5.10). The epoxy serves as a sealant to avoid the seepage of rainwater inside the PL-259 connector. Let the epoxy set and dry.

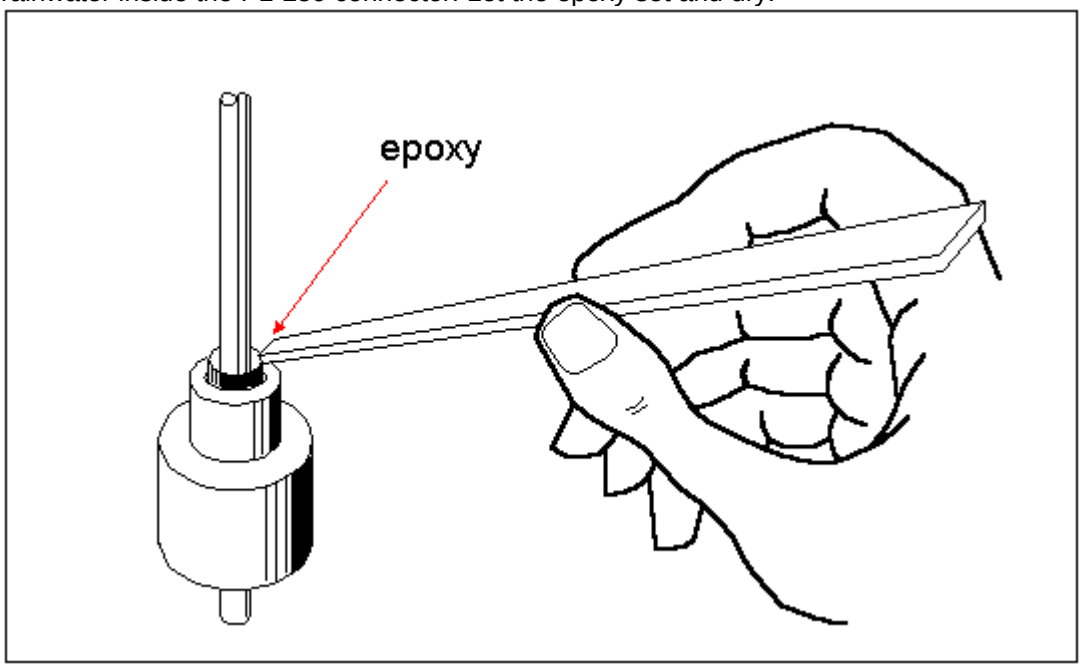


Figure 5.10 Sealing the PL-259 with epoxy glue.

Assembly

First attach the two SO-239 connectors into the two aluminum bushings as shown in Figure 5.11. Don't forget to include its grounding ring or solder lug.

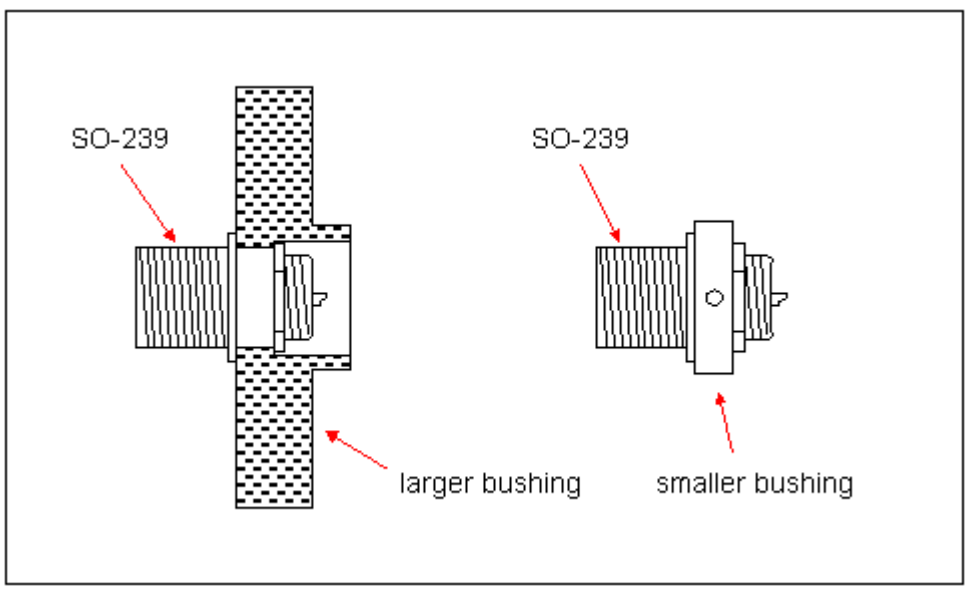


Figure 5.11 Mounting the SO-239 into the larger bushing.

Cut a 36 inches long coaxial cable (RG-58/U) and solder its conductors at one end to the SO-239 connector attached to the larger bushing (see Figure 5.12).

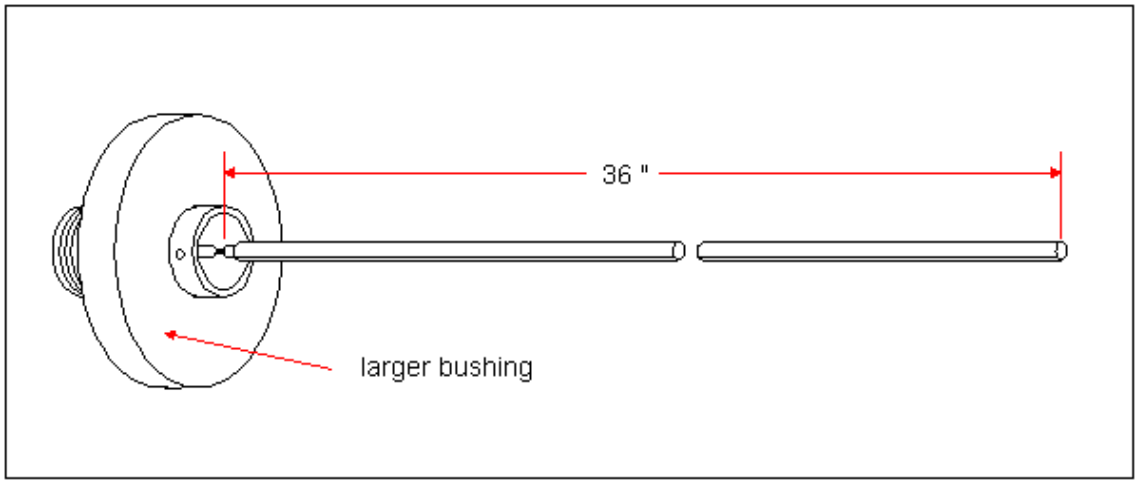


Figure 5.12 Soldering the coaxial cable to the SO-239.

Lay the coaxial cable and the longer tube side by side as they would be when they are finally assembled together. Trim the free end of the coaxial cable at the point 3/8" away from the end of the longer tube (see Figure 5.13).

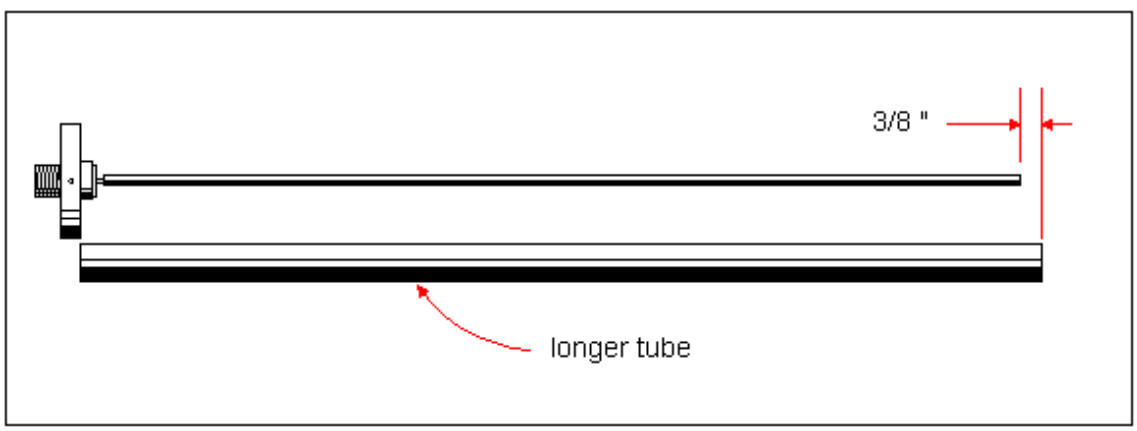


Figure 5.13 Trimming the coaxial cable to the proper length.

Solder the free end of the coax cable into the remaining SO-239 connector attached to the smaller bushing (see Figure 5.14).

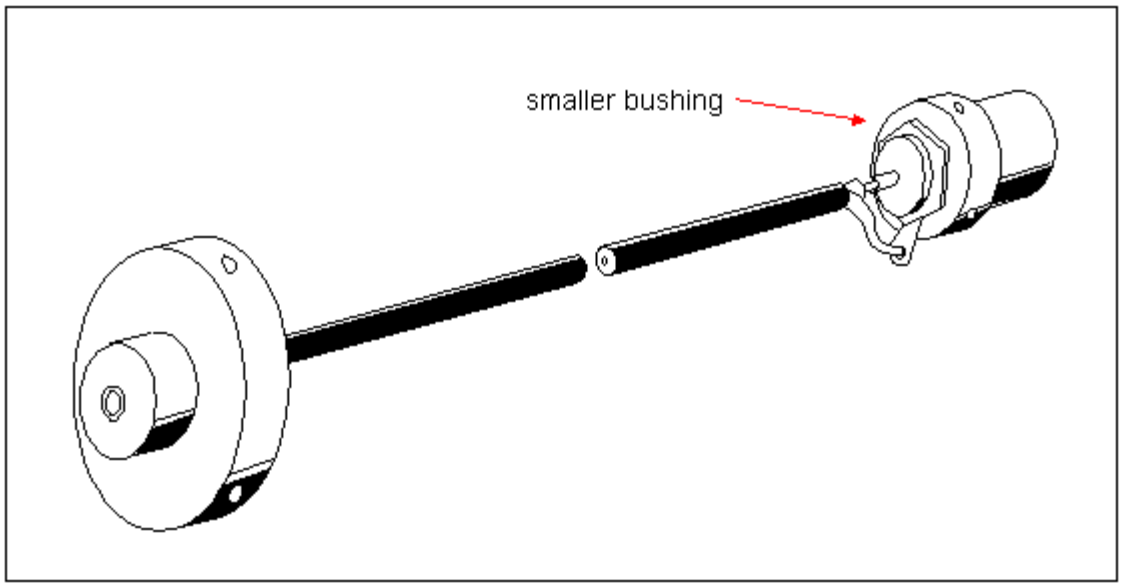


Figure 5.14 Soldering the other end of coaxial cable to the other SO-239.

Insert the smaller bushing, coaxial cable and the large bushing all the way inside the longer tube until the holes in the two bushings are aligned to the holes in the tube itself. If in your first try you did not manage to align the holes then maybe a slight retrimming of the coax cable is needed or the SO-239 connector must be repositioned or resoldered. After a few trials you should have done it right (see Figure 5.15).

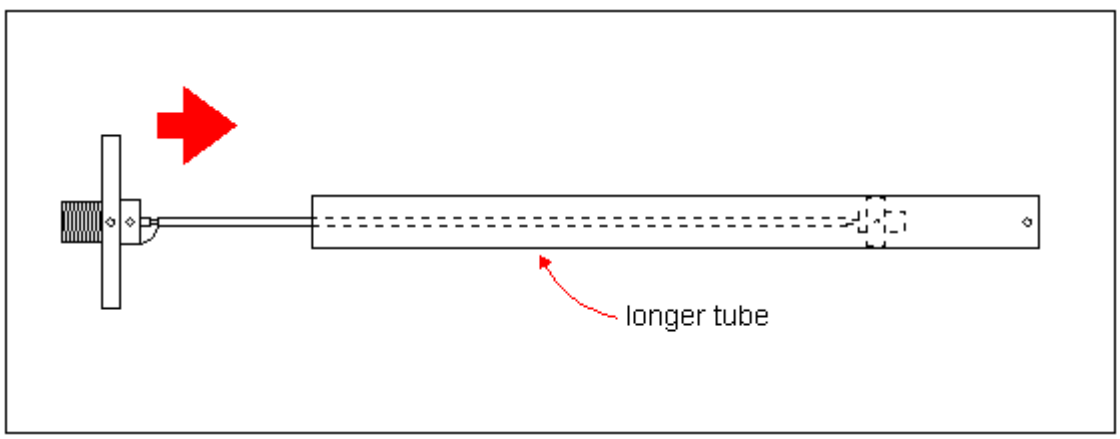


Figure 5.15 Inserting the feeder coaxial cable into the long tube.

Secure the two bushings permanently into the tube using self-tapping metal screws (see Figure 5.16).

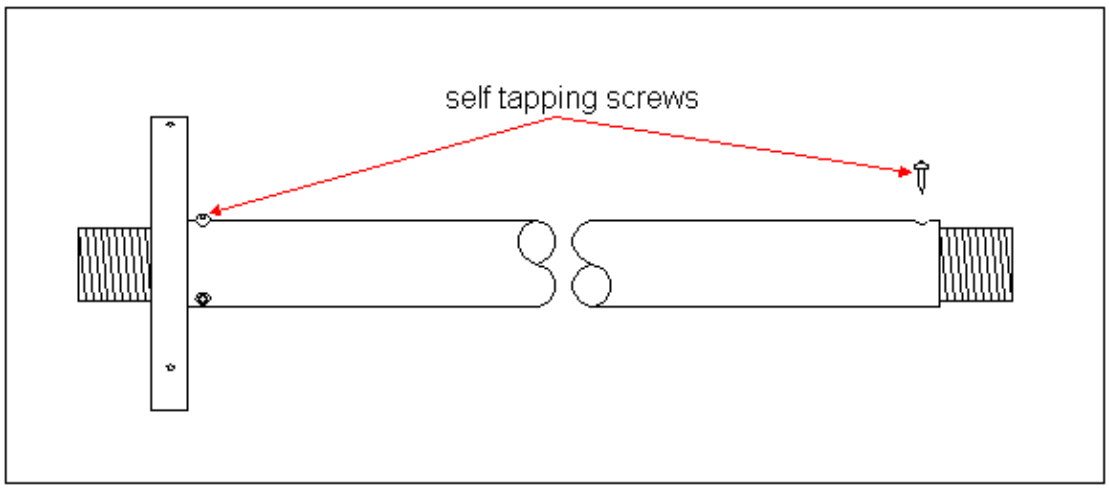


Figure 5.16 Securing the bushings and the tube together.

Insert the longer tube and the large bushing inside the shorter tube (see Figure 5.17). Align the holes in the large bushing to the holes in the shorter tube and place self-tapping screws through the holes to fix the bushing firmly inside the short tube.

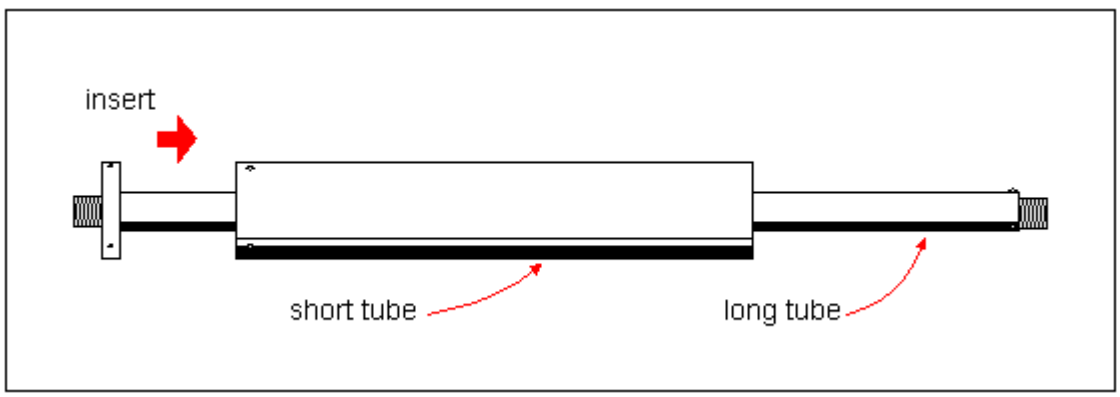


Figure 5.17 Assembling the two tubes together.

Attach the radiator element to the SO-239 connector in the upper larger bushing (see Figure 5.18).

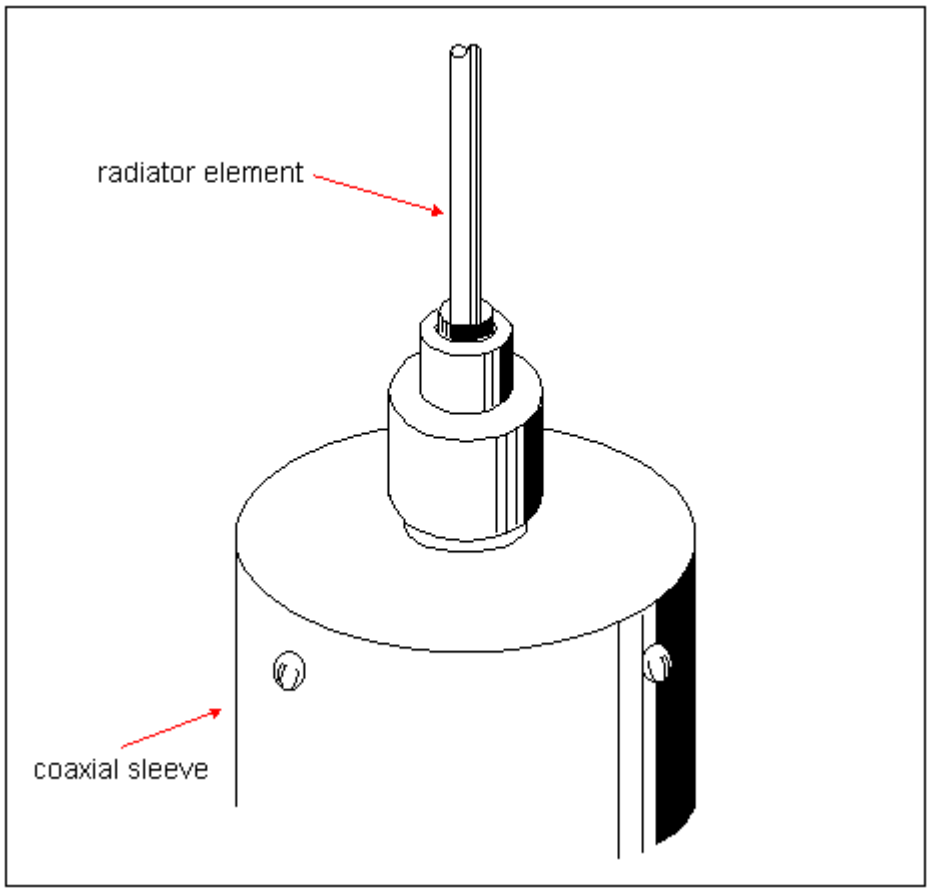


Figure 5.18 Installing the radiator element into the antenna base.

Installation of Model CD-2

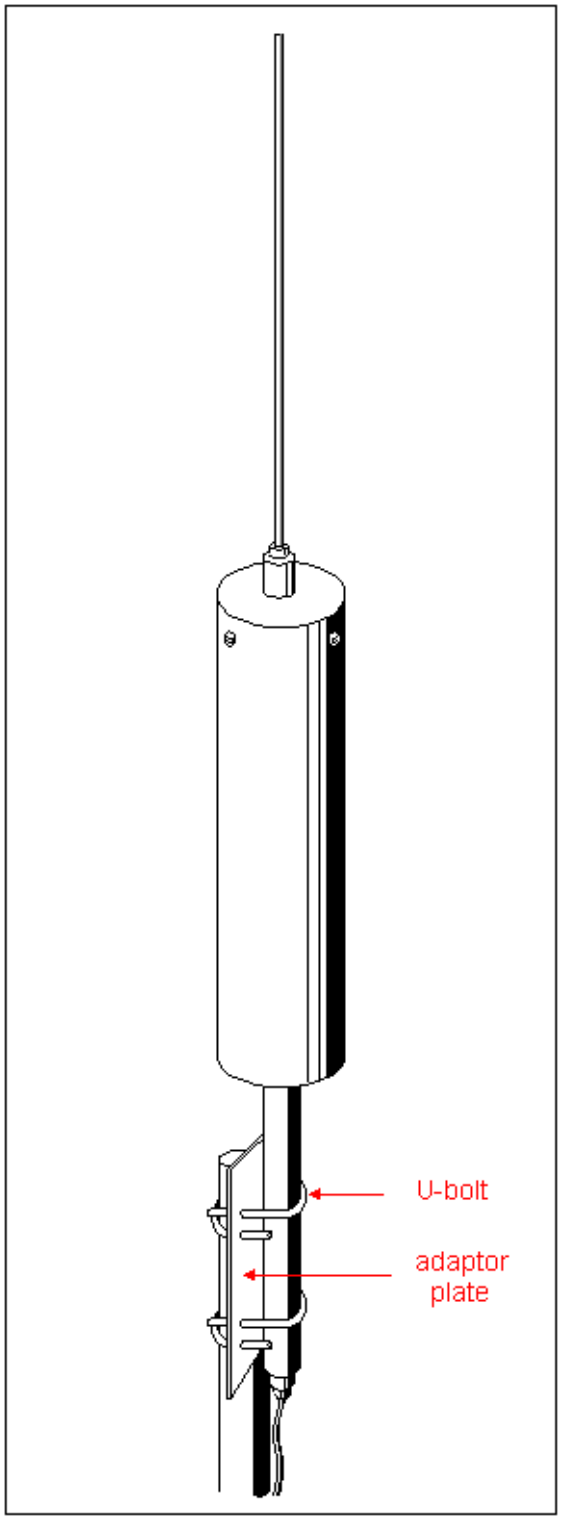
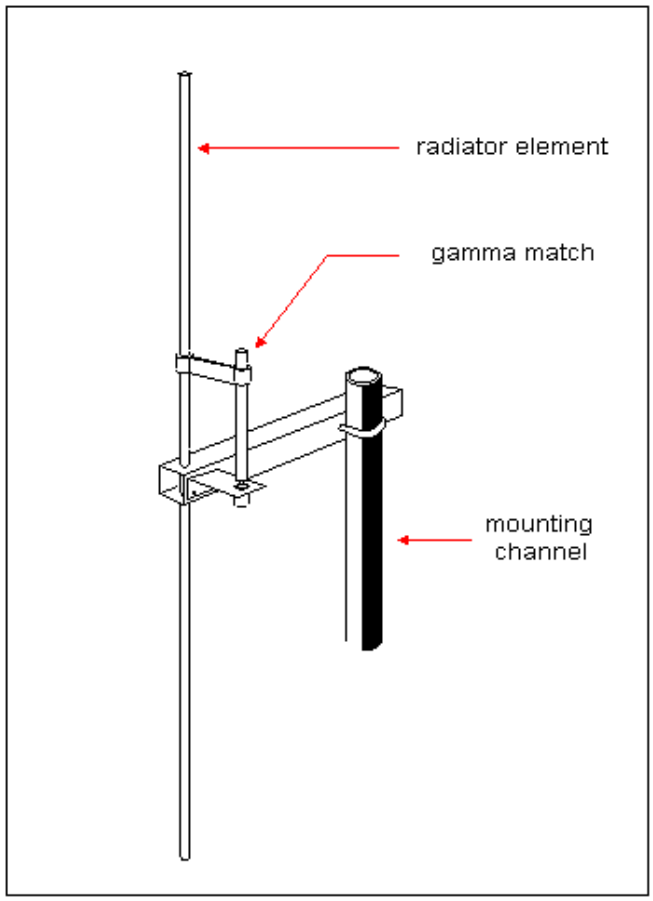


Figure 5.19

6 GROUNDPLANE ANTENNA
(gamma fed)

Model DP-2F

This particular design of a dipole antenna is very popular in VHF applications because of its capability to be fine-tuned during tuning procedures. Tuning is accomplished by a so-called gamma matching system connected near the center of the dipole element. Gamma matching is based on the principle of delta match system where the transmission line can be directly connected near the center of a continuous halfwave conductor and fanned out and tapped at the point of most efficient power transfer.



The middle of a halfwave dipole is electrically neutral - meaning there is no RF voltage present so the outer conductor of the coax cable can be connected directly to the element at this point. The inner conductor of the coaxial cable carries an RF current so it is tapped into the dipole element at the matching point.

After a careful observation of this design you will notice that the center conductor of the coax cable is not directly connected to the dipole element but instead coupled via a short tube called "gamma tube". The combination of the short tube and the coaxial cable inside it provides the capacitance needed to cancel the inductance of the dipole element to attain an electrical balance. The gamma match therefore achieves two functions at the same time, that is to match the impedance of the transmission line to the impedance of the antenna and to couple the unbalanced coaxial cable to the symmetrical dipole element. This method makes it unnecessary to use a separate balancing transformer. Fine tuning of the antenna can be done by adjusting the shorting bar that connects the gamma tube to the dipole element until the lowest SWR response is achieved.

Model DP-2 is also used as a basic driven element for high gain Yagi and collinear antenna designs. Because the middle of the dipole element is electrically inactive, it does not require to be insulated from its mounting boom thereby simplifying the mechanical construction. Lightning protection for this antenna system is also improved because all the metallic parts of the antenna are grounded via its mast or tower.

The dipole design described in this chapter is designed to operate in the frequency band of 140-150 MHz. If properly tuned it exhibits an SWR of less than 1.4:1 over the entire band. It radiates its signal in an omni-directional pattern. It has a gain of 1 dB (unity gain) compared to a standard dipole reference.
This antenna is intended primarily for fixed installations. However, some radio operators were able to use it succesfully in mobile operations by modifying its mechanical construction.

Some antenna constructors choose to build this antenna because it presents them deeper understanding of the electrical principles of antennas compared to other simpler designs like groundplanes or coaxial dipoles. If you are the experimenter type of radio operator then this design is for you.

Materials List

Quantity	Specification/Description	Dimensions
1	Aluminum tube 3/8" id*	38" long
1	Aluminum tube 3/8" id*	6" long
1	Aluminum square channel	1" x 1" x 12"
1	Aluminum strip - see text for fabrication	1/2" x 4"
1	Coax cable RG-58/U	6"
1	BNC female connector	
2	Stove bolts - brass or GI	1/8" x 3/8"
2	Hex nuts - brass or GI	1/8" id
1	U-bolt with accompanying hex nuts and washers	
3	Self tapping metal screws	1/8" x 1/2"
Miscellaneous:	Epoxy glue	

*id- inside diameter

Construction

The radiator element is made from 3/8 od* aluminum tube cut to a length of 38 inches. Drill a hole (1/16" diameter) through and through at the middle of its length (see Figure 6.2).

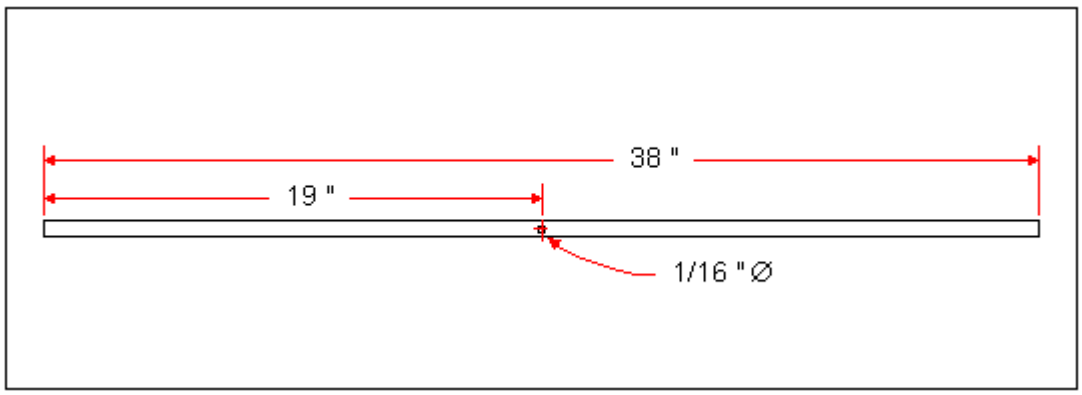


Figure 6.2 Drilling a hole through the middle point of the radiator element.

Next prepare the mounting channel by drilling a hole at one end (see Figure 6.3). The diameter of the hole must accomodate the aluminum tube that will be inserted into it. The hole is 3/8" and slightly oversized so that the tube will not be scratched upon insertion but not too loose as to sacrifice rigidity.

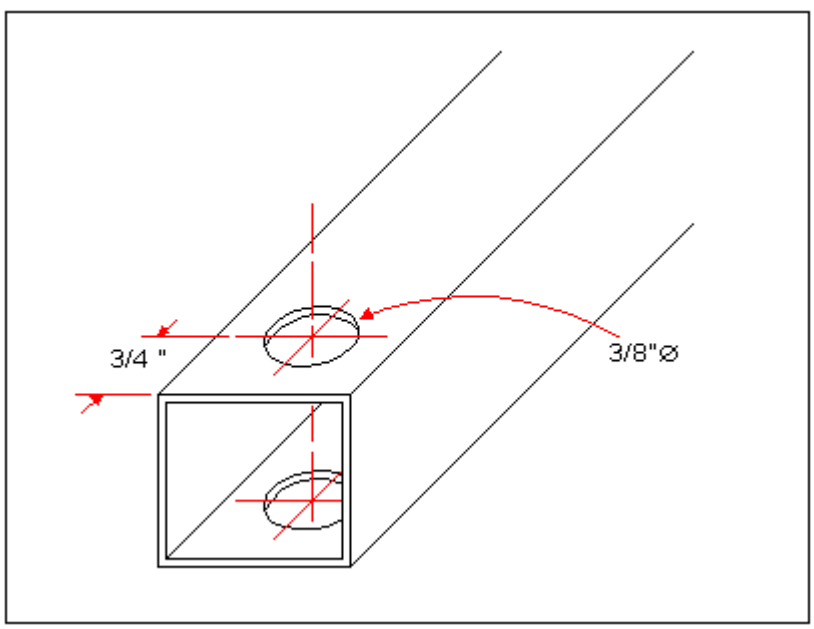


Figure 6.3 Preparing the mounting channel.

Next, drill two small holes (1/16" diameter) at one side of the channel perpendicular to the axis of the bigger hole (see Figure 6.4).

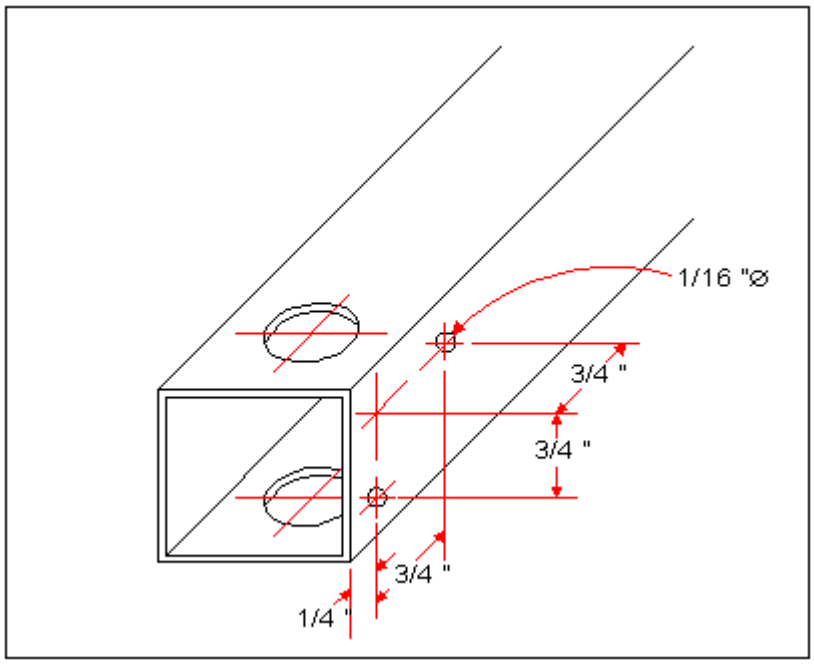


Figure 6.4 Drilling holes for the gamma mounting bracket.

Drill another pair of holes (3/16" diameter) at the same side but at the opposite end of the channel (see Figure 6.5). Drill the hole through and through. The size of the holes and the distance between them must conform to the dimensions of the U-bolt used.

Figure 6.5 Drilling holes for the U-bolt at the opposite end .

Drill a single 1/8" diameter hole at the other side opposite to the two small holes (1/16" diameter). See dimensions in the following illustration.

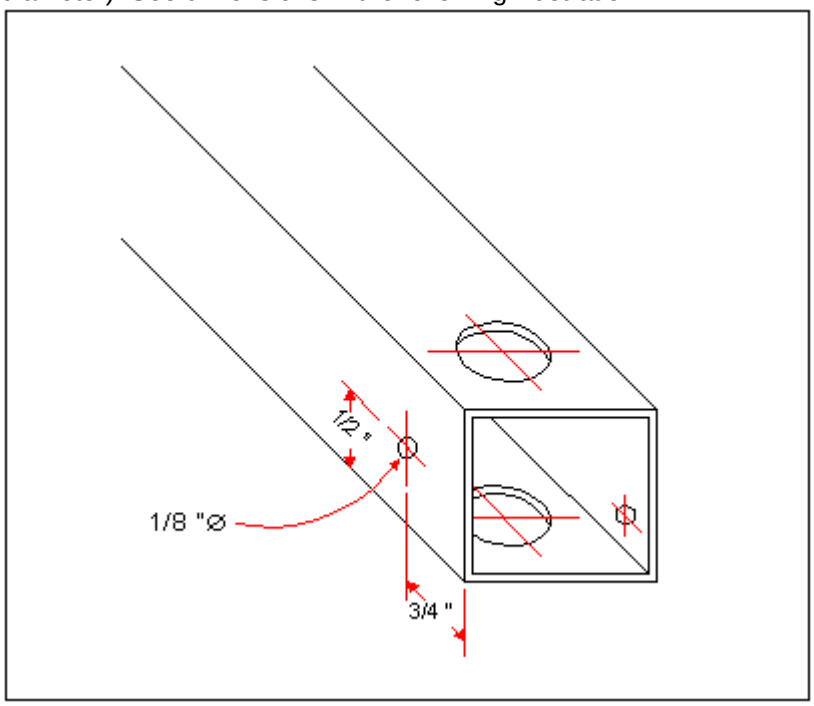


Figure 6.6 Drilling a single hole.

Insert the aluminum tube through the large hole and align the hole at its middle part to the 1/8" diameter hole at the side of the channel (see Figure 6.7).

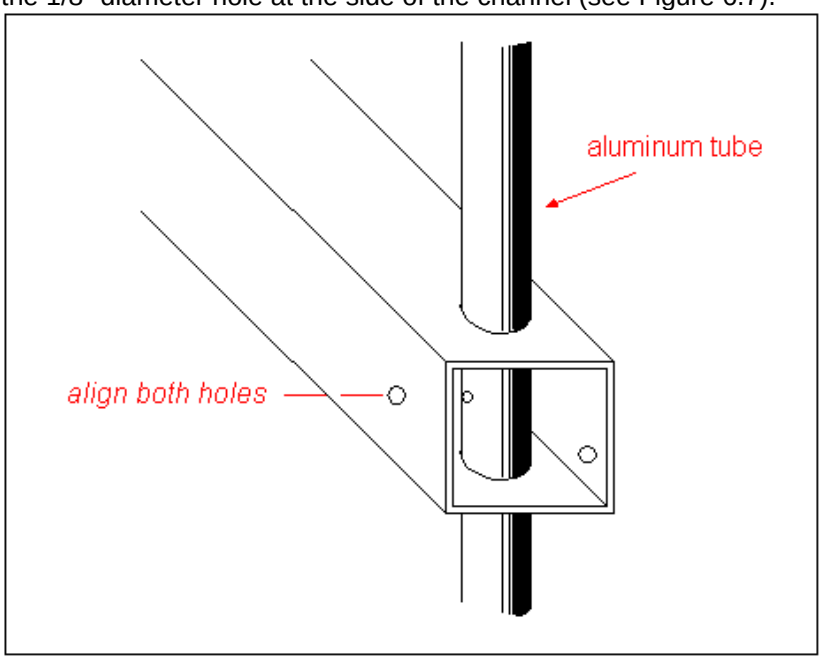


Figure 6.7 Inserting the aluminum tube into the mounting channel.

Insert a self tapping screw through the sidehole and forcibly screw it into the smaller hole of the tube inside (see Figure 6.8). Tighten the screw until the aluminum tube is rigidly held in the aluminum channel.

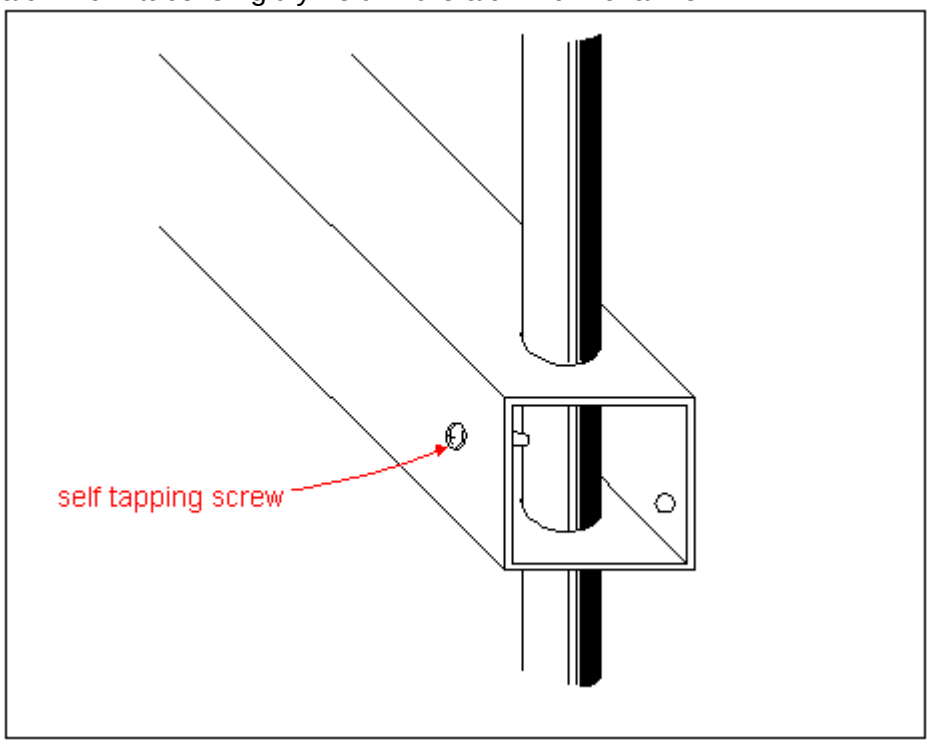


Figure 6.8 Locking the tube with a self tapping screw.

Prepare the feedpoint angle bracket. The bracket is cut from a small strip of aluminum and bent into a right angle. An alternative method is to saw off a portion of a 1" x 3" rectangular aluminum channel. This will give you a more durable bracket with a near perfect angle.

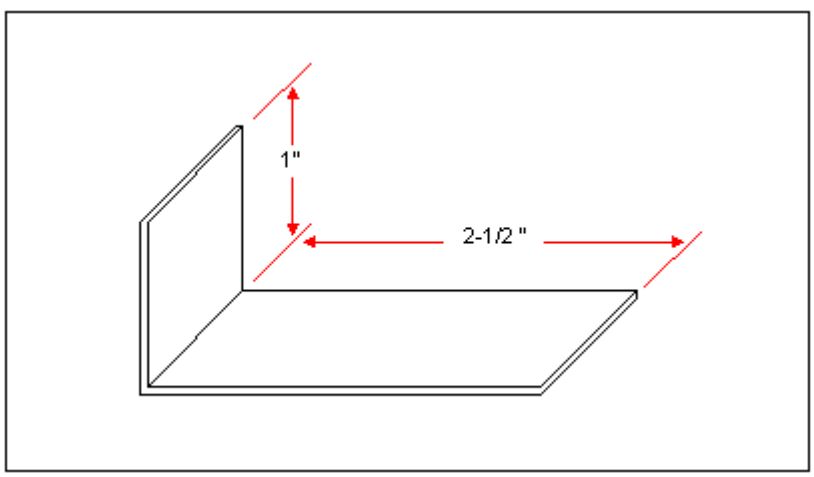


Figure 6.9 Preparing the feedpoint bracket (gamma mounting bracket).

Drill two small holes (1/8" diameter) at one side of the angle bracket. Drill another hole at the other side of the bracket. This lone hole must be large enough to accomodate the BNC female connector (see Figure 6.10).

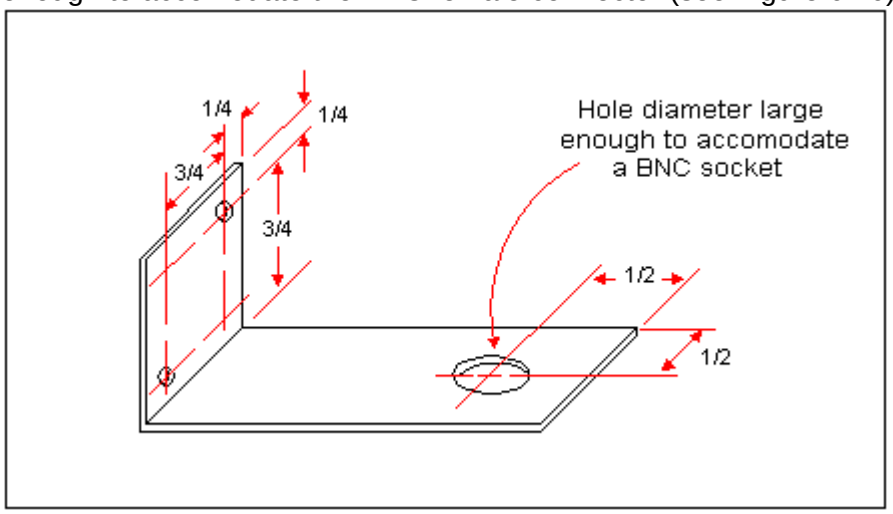


Figure 6.10 Drilling mounting holes in the bracket.

Attach the bracket into the mounting channel by screwing it with small self tapping screws (see Figure 6.11).

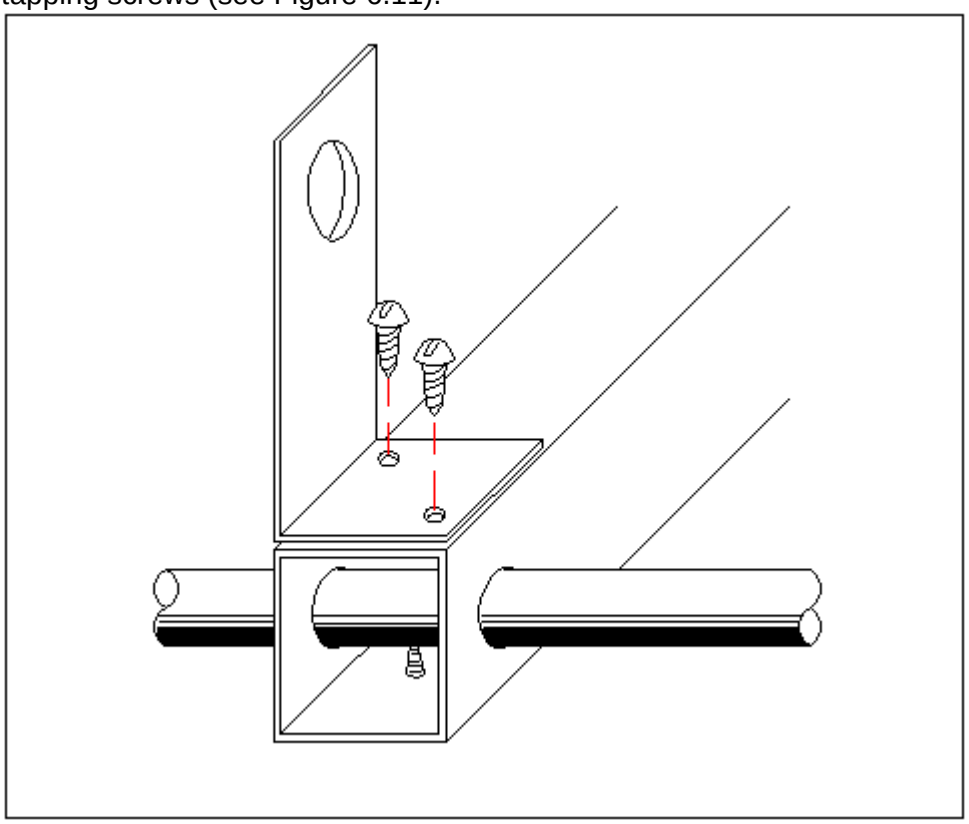


Figure 6.11 Fixing the feedpoint bracket on the mounting channel.

Insert the BNC female connector in an upside down position into the large hole of the feedpoint bracket and secure it with its nut (see Figure 6.12).

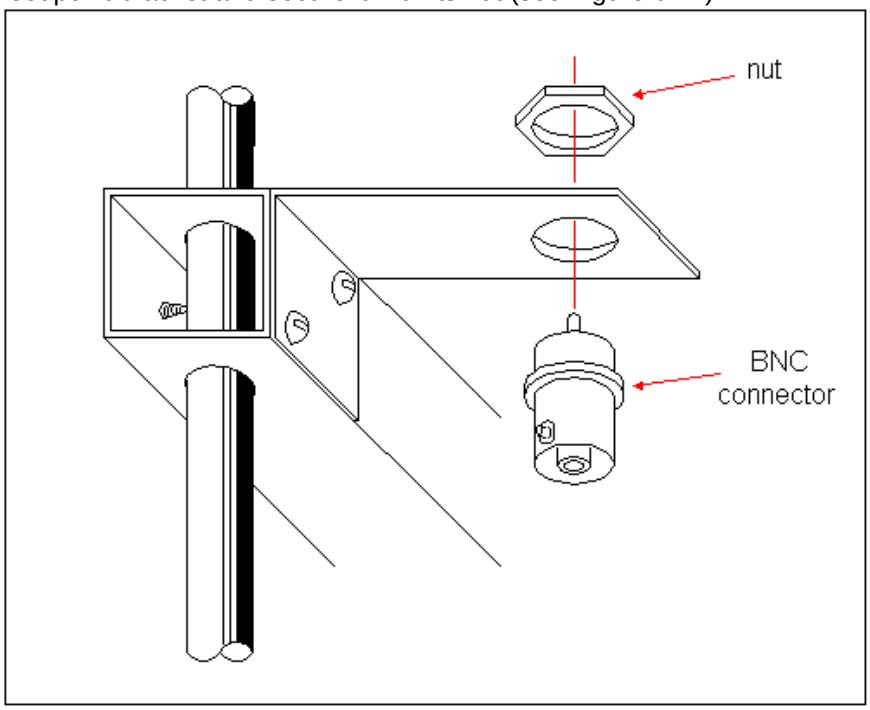


Figure 6.12 Installing the BNC connector into the feedpoint bracket.

Next step is to prepare the tuning clamp and the gamma matching tube. First, fabricate a flat strip from a scrap tube (about 4 inches long) by pressing it in a table vise until it is completely flattened. Cut about 4 inches of the flat strip and form both ends to a ring clamp by bending it around an aluminum tube. The ends must be formed to fit around the tube (see Figure 6.13).

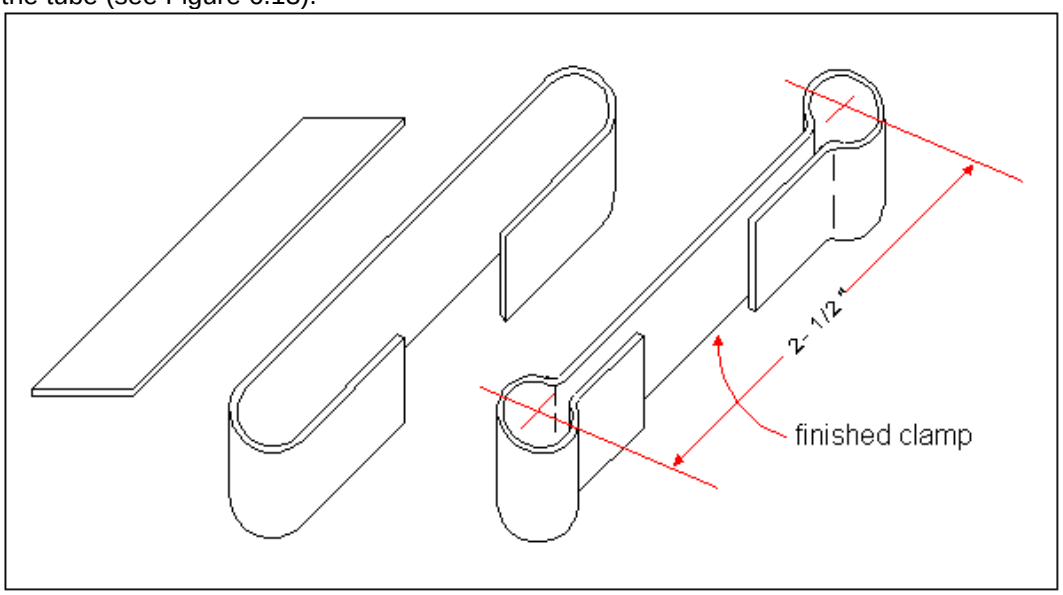


Figure 6.13 Preparing the tuning clamp.

Drill two holes (1/8" diameter) in the tuning clamp (see Figure 6.14).

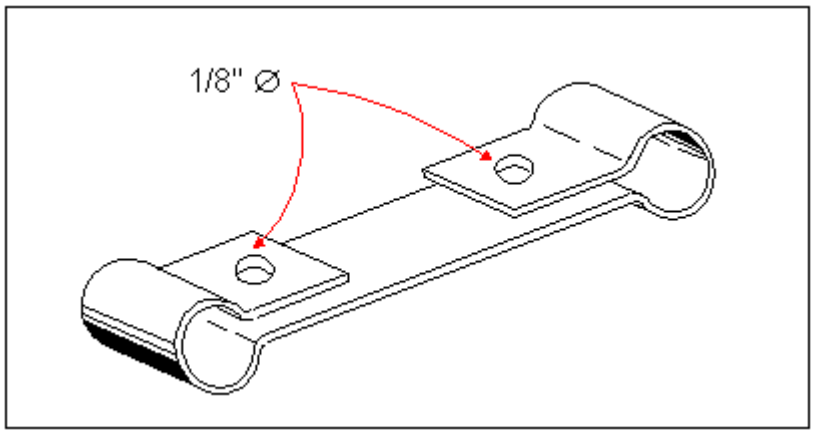


Figure 6.14 Drilling holes in the tuning clamp.

Insert the 6 inch long tube into one loop of the clamp and secure the clamp with a 1/8" x 3/8" stove bolt. Attach a nut to the bolt and tighten it lightly (see Figure 6.15). **Don't tighten this bolt too much at this time!**

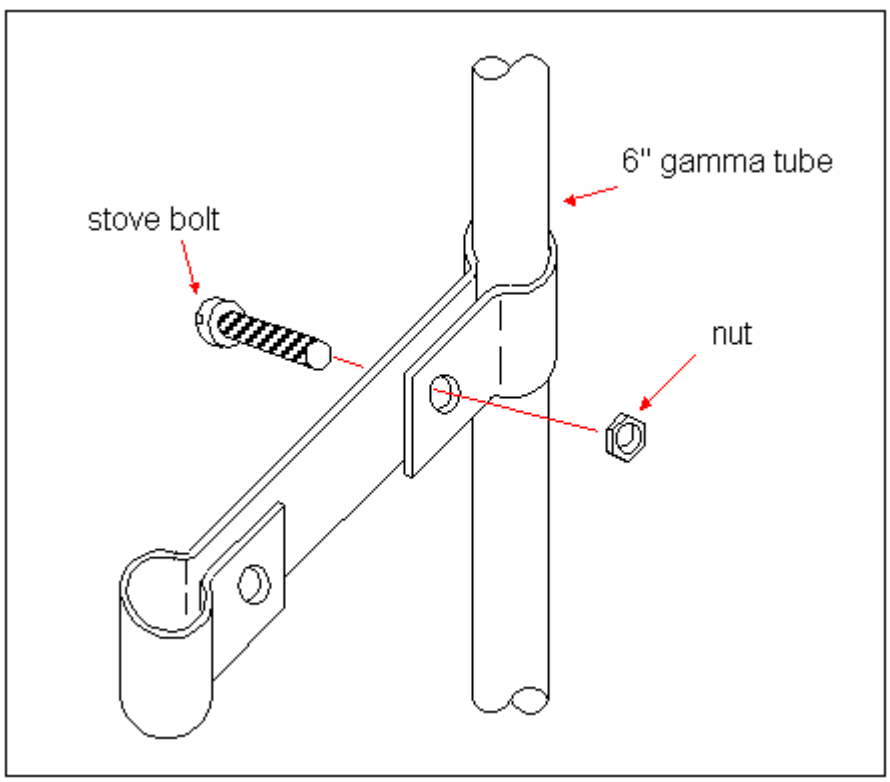


Figure 6.15 Inserting the 6 inch long gamma tube into the tuning clamp.

Waterproof the top end of the 6 inch gamma tube by inserting a substantial volume of cotton wad inside the open end. The cotton wad serves as a stopper for the epoxy. Place epoxy glue over the cotton wad inside. Let the epoxy set and dry (see Figure 6.16).

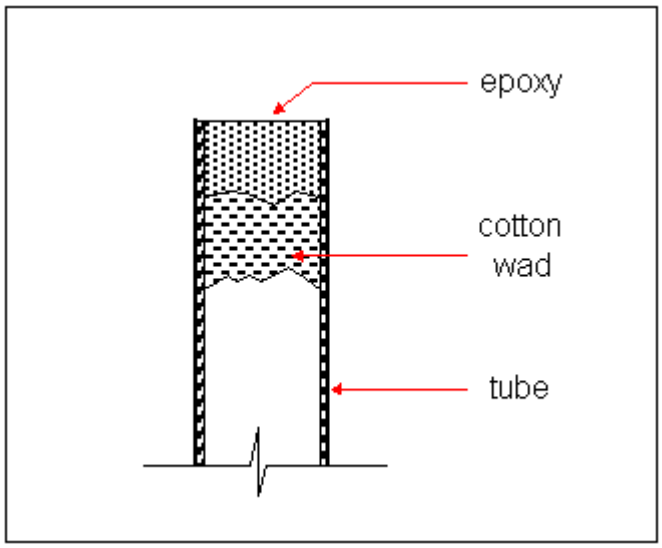


Figure 6.16 Sealing the top end of the gamma tube.

While the epoxy is drying, prepare the gamma match from a short length of **RG-8/U** coaxial cable. Cut a piece 6 inches long and remove its vinyl outer jacket and its braid. Cut away a small portion of the PE inner insulator exposing the copper conductor inside (see Figure 6.17).

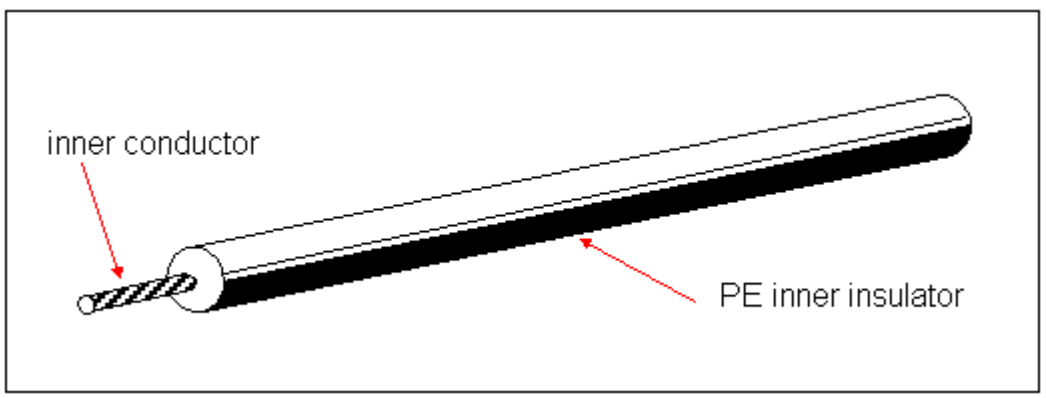


Figure 6.17 Preparing the gamma match from a short length of RG-8/U.

Solder the exposed copper conductor of the gamma match directly into the center pin of the BNC connector attached to the mounting channel.

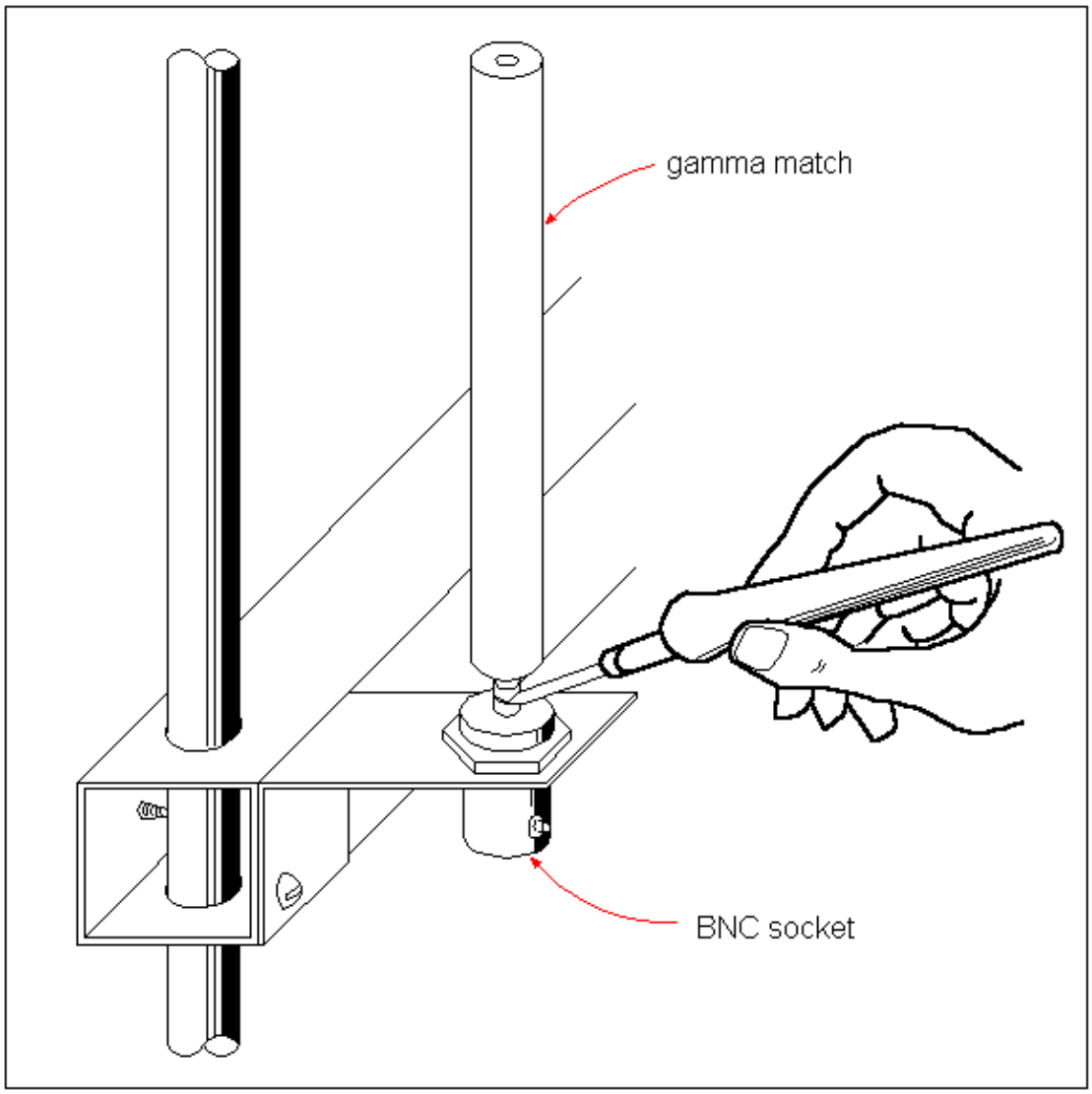


Figure 6.18 Soldering the gamma match into the BNC connector.

Next, carefully insert the free end of the tuning clamp into the radiator element (long tube) starting at the top end. As the tuning clamp is lowered down along with the gamma tube attached to it, insert the gamma match into the gamma tube (see Figure 6.19).

Stop the gamma tube just about 1/2 inch above plane of the angle bracket or just enough for the gamma tube to cover the whole PE insulator. Insert a bolt into the tuning clamp attached around the radiator element and tighten it lightly enough to hold the gamma matching assembly in place. The antenna is already mechanically ready at this point, it only needs to be tuned to resonance for proper operation.

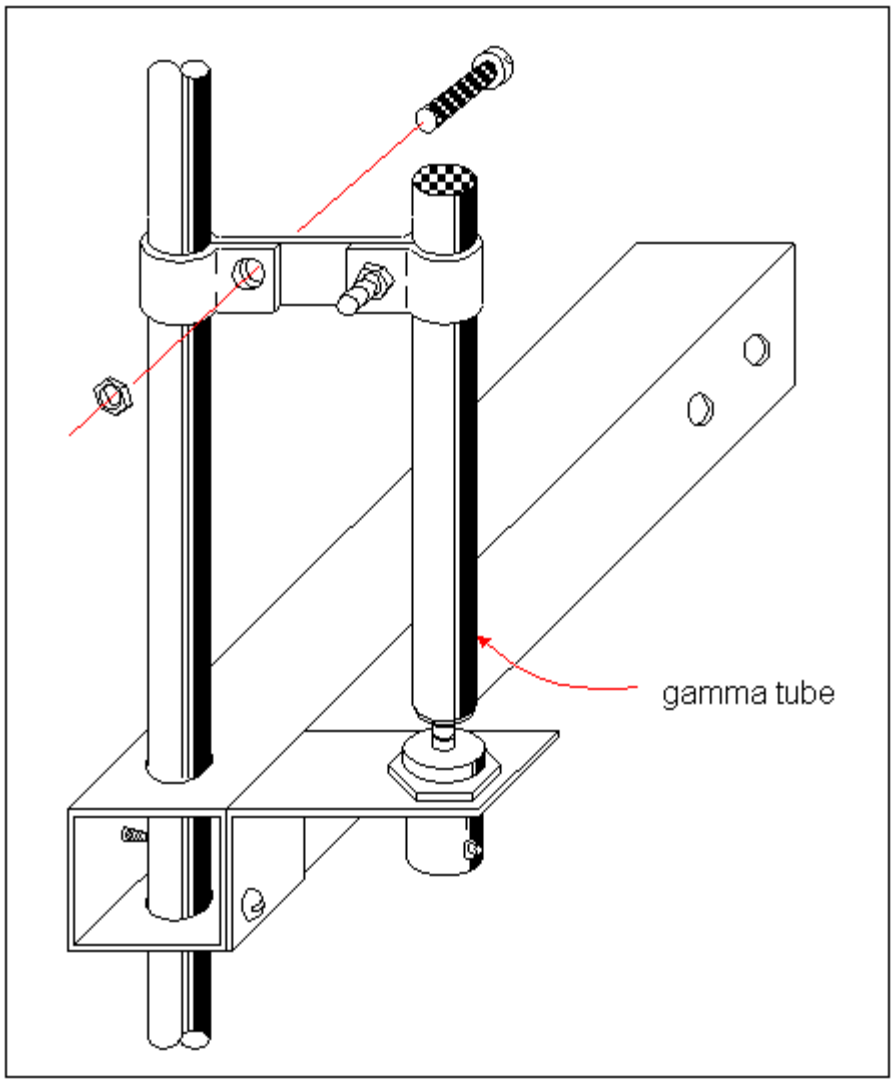


Figure 6.19 Assembling the gamma tube into the gamma match.

In tuning the antenna to resonance, install it to the mast or tower following Figure 6.1. Connect the coaxial cable to the BNC connector in the angle bracket and connect the other end to a suitable SWR meter. Connect a VHF transceiver to the SWR meter (usually marked 'transmitter') and set its frequency to either the center or extreme frequencies of the band. Key the PTT to transmit and read the SWR response.

To tune the antenna: move the tuning clamp and find the position where you can get the lowest SWR response for the center frequency and a relatively flat response curve over the entire band. Move the tuning clamp either lower or higher about 1/4 inch at a time (see Figure 6.20).

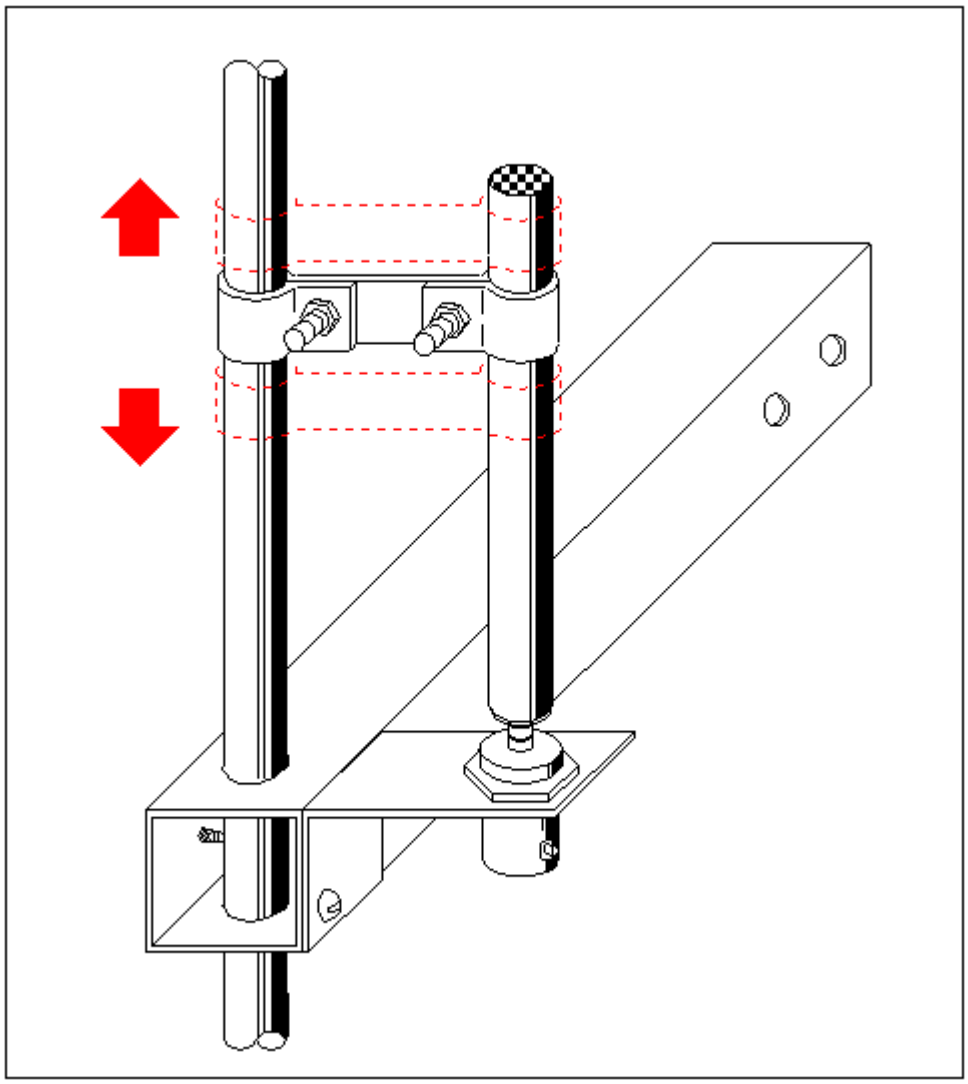


Figure 6.20 Raising or lowering the tuning clamp to find the best match.

If you have finally found the right position (after several trials) then you have succesfully tuned the antenna to resonance. Tighten the nuts at the tuning clamp permanently and place a moderate amout of silicone sealant (RTV compound) around the open lower end of the gamma matching tube to seal it off from moisture and rainwater (see Figure 6.21).

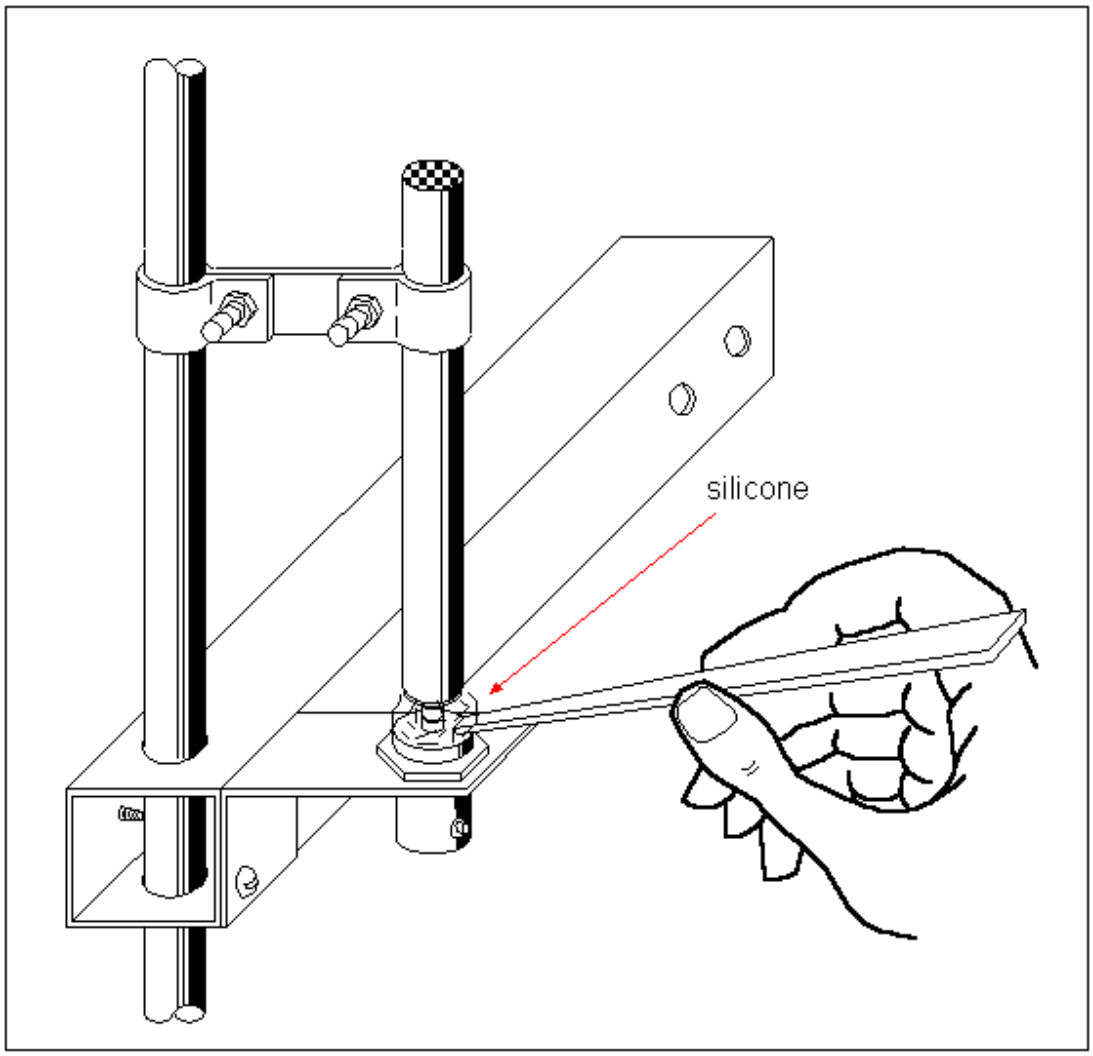


Figure 6.21 Sealing off the lower end of gamma tube with silicone.

IMPORTANT:

- **Do not touch any part of the antenna while keying the rig or reading the SWR response.**
- **Do not move the tuning clamp while the transceiver is still transmitting.**

7 QUADLOOP ANTENNA

Model QA-2F

In preceeding chapters all of the various antenna designs presented are assemblies of linear halfwave (or approximately halfwave) dipole elements. On the other hand other element forms may also be used to effectively function as an antenna. One example is the quad antenna described in this chapter. This is the type of antenna with a radiating element made of a loop having a perimeter of one wavelength and used in much the same way as a dipole.

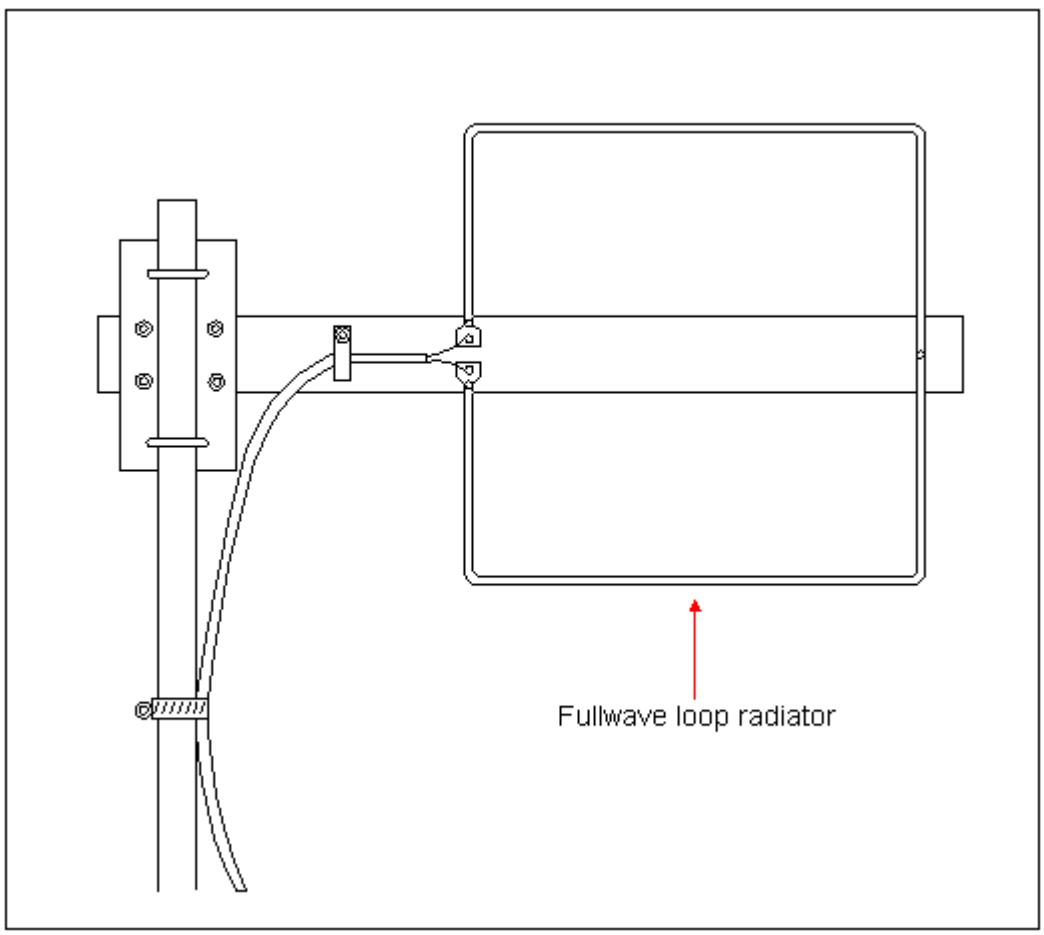


Figure 7.1 A quad loop antenna model QA-2F.

The quad antenna was originally designed in the late 1940's. Since then it has been the subject of controversy whether it performs better than a dipole. The debate continues but after some years several facts have become apparent. It was found out that the quad has a slight gain of approximately 2 dB over a dipole. It is also said to cover a wider area in the vertical plane and exhibits broadband characteristics.

The quad antenna model QA-2F is specifically designed to operate in the frequency band of 140-150 MHz. It displays a bi-directional radiation pattern with maximum radiation in the direction perpendicular to the plane of the loop. By carefully following the instruction for constructing this antenna you should be able to get an SWR response of less than 1.5:1 over the entire band.

Materials List

Quantity	Specification/Description	Dimensions
1	Aluminum tube 3/8" id*	82 " long
1	Plastic plate 1/2 " thick	32" long
2	Stove bolts - brass or GI	1/8" x 3/8"
3	Stove bolts - brass or GI	1/8" x 1"
4	Stove bolts - brass or GI	3/16" x 1"
2	U-bolts with accompanying hex nuts and lockwashers	
1	Plastic C-clamp - enough to hold a 3/8" cable	
1	Self tapping metal screw	1/8" x 3/8"
2	Eye terminals - vinyl insulated	
4	Plain washers - 1/8" id*	
1	Hose clamp - enough to hold a 1-1/2" tube	

*id- inside diameter

Construction

First, prepare the plastic mount with dimensions shown in Figure 7.2.

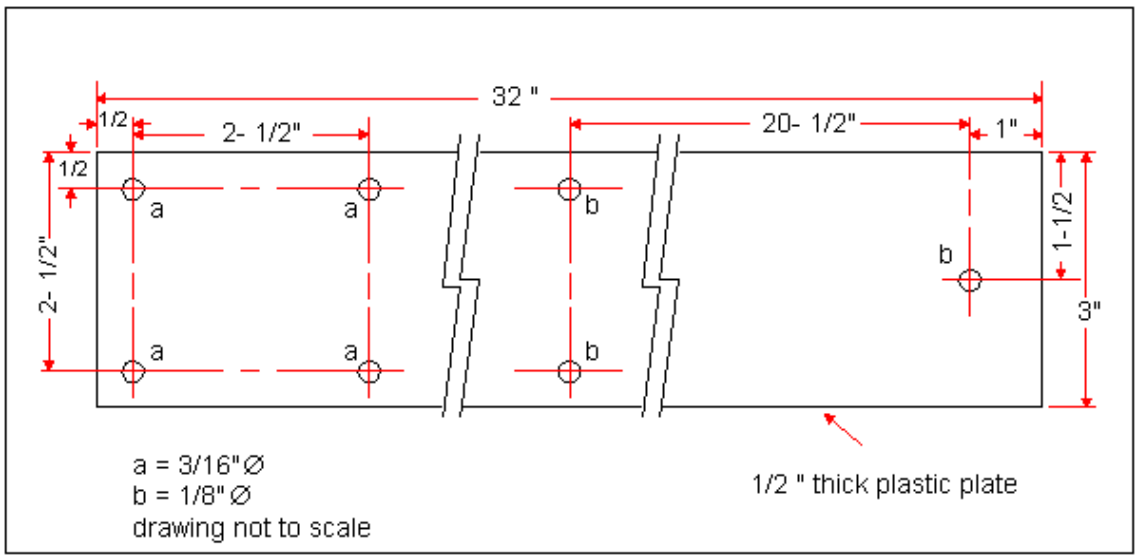


Figure 7.2 Plastic mount dimensions.

Next, prepare the metallic mast adaptor. As shown in the following illustration, the distance between one pair of 3/16" holes at the extreme ends is equal to the distance between the threaded ends of the U-bolt used (see Figure 7.3).

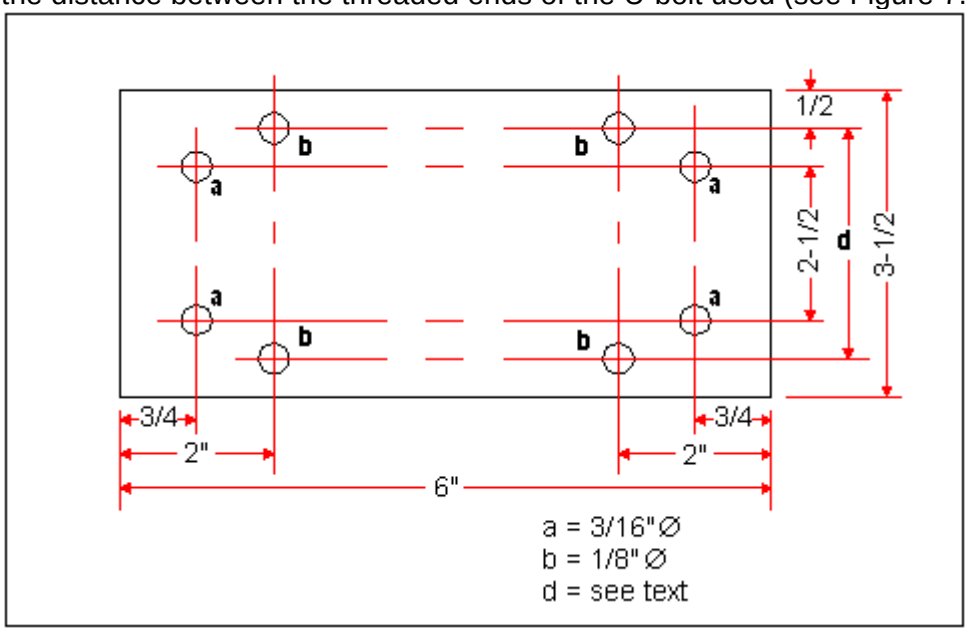


Figure 7.3 Mast adaptor dimensions.

Join the two plates together using four 3/16" x 1" stove bolts made of rust resistant materials (e.g. brass or stainless steel). Do not forget to include a lockwasher in each bolt (see Figure 7.4).

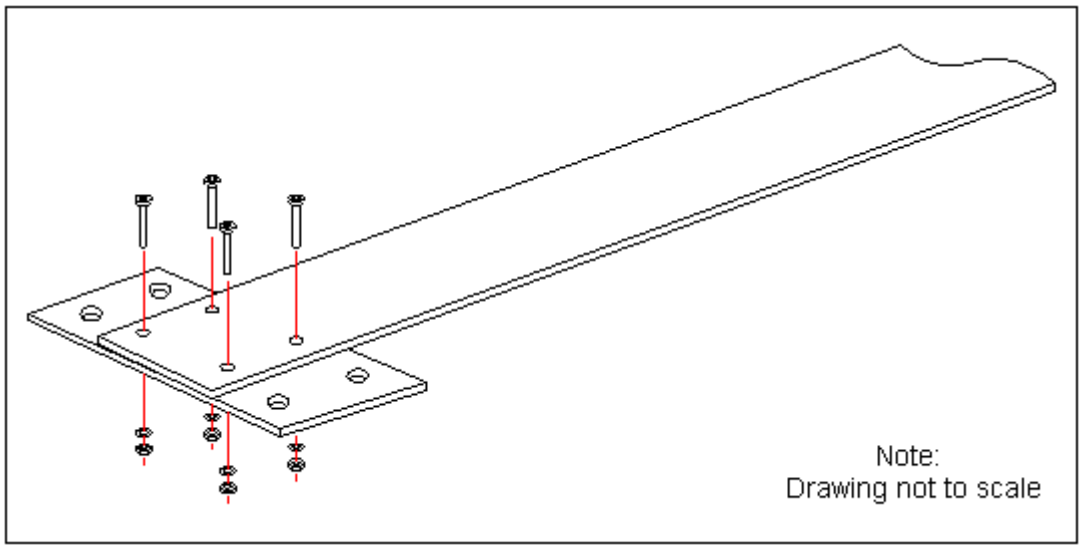


Figure 7.4 Joining the two plates together with stove bolts.

Bend the aluminum tube into a square loop with equal sides using a suitable tube bender (see Figure 7.5). Cut away the excess tube.

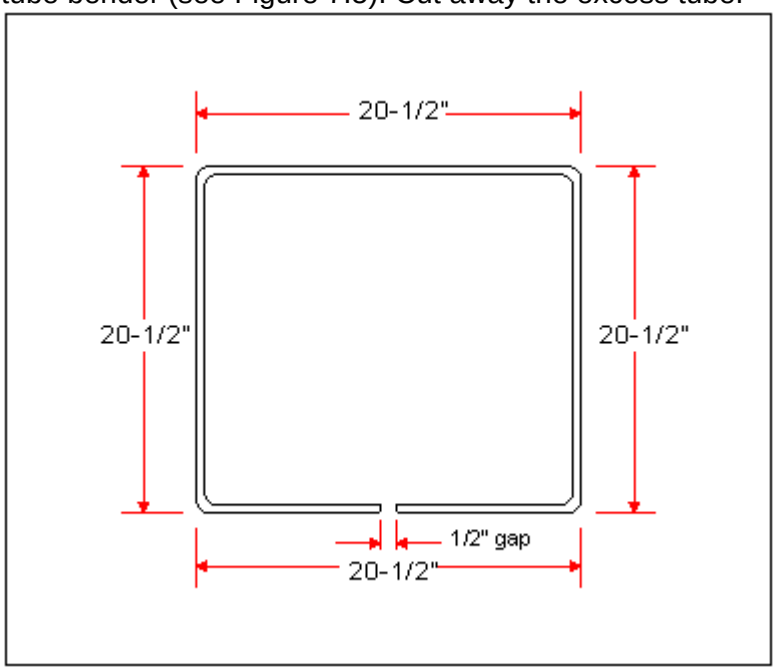


Figure 7.5 Forming the tube into a square loop.

Flatten a small portion at both ends of the tube and drill a hole (1/8" diameter) in each flattened end (see Figure 7.6).

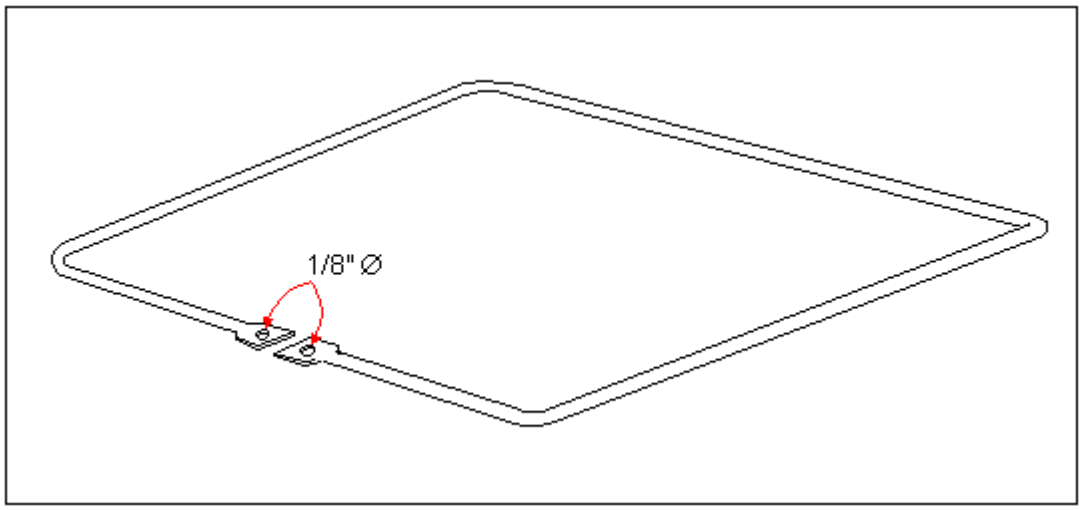


Figure 7.6 Flattening and drilling the ends of the tube.

Drill additional holes (1/8" diameter) in the tube as shown in the following illustration (Figure 7.7). The holes must be drilled through and through. Be careful in drilling the holes to avoid deforming the tube.

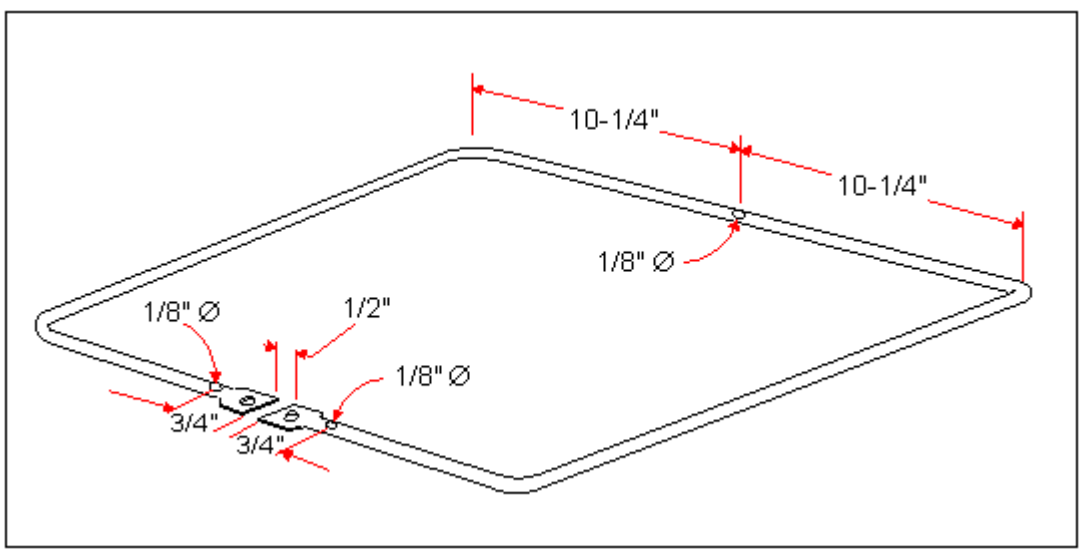


Figure 7.7 Drilling additional holes in the tube.

Insert two stove bolts (1/8" x 3/8") through the holes at both ends of the loop and attach the necessary hardware as shown in Figure 7.8.

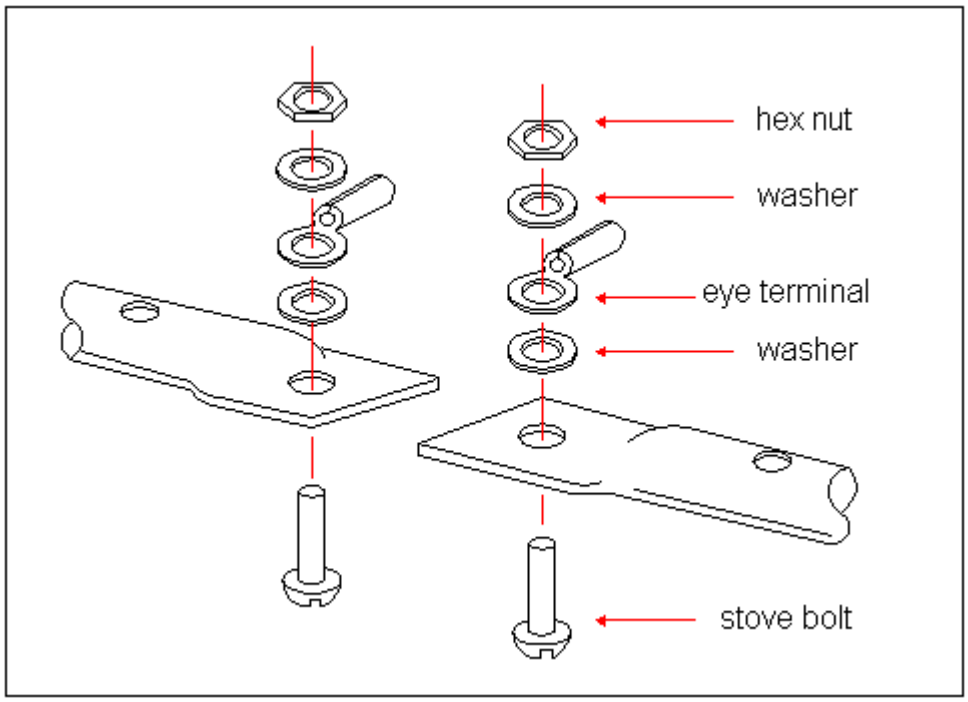


Figure 7.8 Installing the necessary hardware at both ends of the tube.

Attach the prepared loop into the plastic mounting plate by bolting it through the 1/8" diameter holes as shown in the following illustration. Use 1/8" x 1" stove bolts (brass or stainless steel). See Figure 7.9.

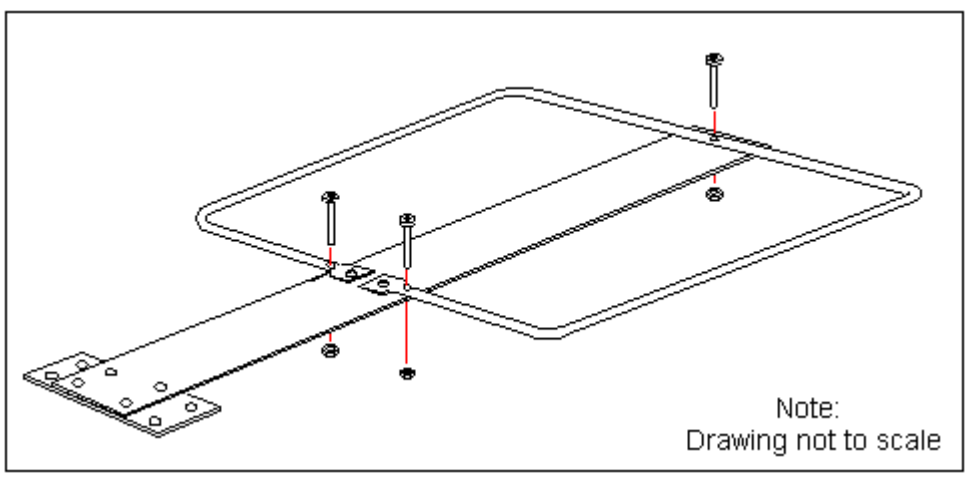


Figure 7.9 Fixing the loop on the plastic mount.

Prepare one end of the coax cable by separating the inner conductor from the copper braid. Solder the two conductors to the two eye terminals in the loop. The braid is costumarily connected to the lower terminal (see Figure 7.10).

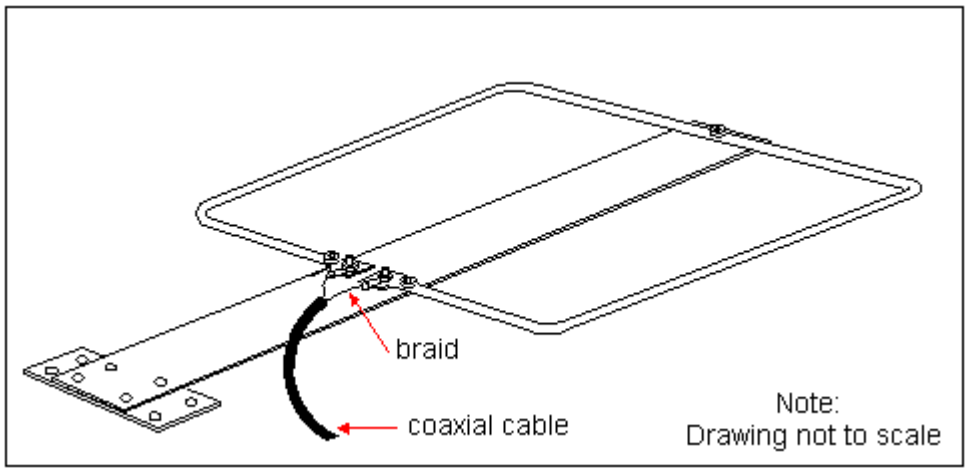


Figure 7.10 Connecting the coaxial cable to the loop element.

Clamp the coaxial cable to the plastic mounting plate (see Figure 7.11).

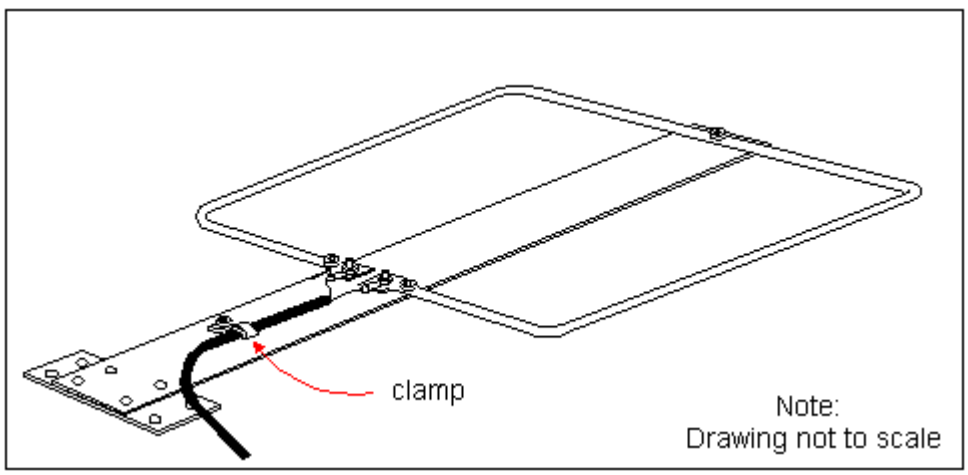


Figure 7.11 Clamping the coaxial cable to the plastic mount.

Installation of QA-2F

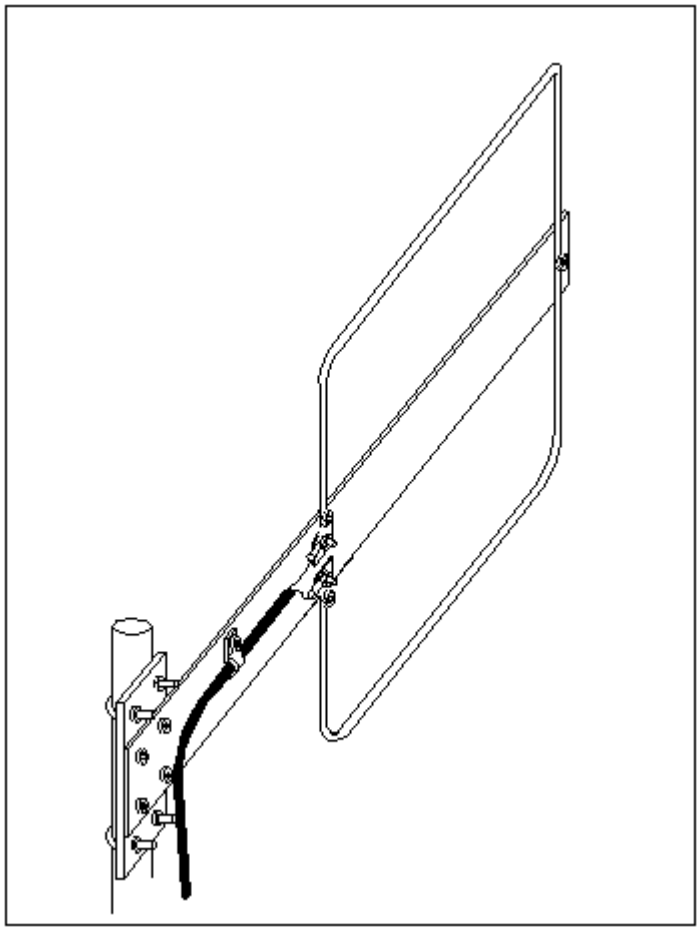


Figure 7.12 Installing the QA-2F to the mast.

8 DISCONE ANTENNA

Model CD-2W

Most of the antenna designs described in the preceeding chapters are all suitable for VHF work requiring omni-directional pattern of radiation. Also in the mechanical viewpoint, these designs are simple and easy to construct which makes them very popular among radio operators. However all of them have a limited bandwidth of 140-150 MHz. If one attempts to operate his transceiver outside these frequency limits (assuming he has a wideband transceiver) the signal response becomes weaker as the operating frequency of the transceiver is moved farther away from the operational bandwidth of the antenna. At the same time the SWR in the transmission line increases and can reach an intolerable point which may cause damage to the transceiver. Although this handicap can be avoided by using a different antenna tuned to a different frequency band, the process of changing antennas everytime the operator changes his operating band becomes time-consuming and cumbersome. This problem can be solved by using a discone antenna described in this chapter.

The discone antenna is a broadband antenna. Meaning it can operate over a wide range of frequencies. Theoritically, a properly designed discone antenna can operate up to a frequency 10 times the value of its lowest operational frequency. Specifically speaking, if a discone antenna is designed to operate with a lowest operational frequency of 140 MHz, then it can be conveniently used up to 1.4 Gigahertz! The lowest operational frequency is called cut-off frequency. Below this frequency the SWR will increase rapidly.

Amazing! Well, a discone antenna can achieve that because it functions more like a transformer than a conventional antenna. It couples the low impedance transmission line to the higher impedance of free space. Its signal pattern is similar to that of a quarterwave groundplane antenna. Radiowaves from the transmission line emerge at the feedpoint (cone apex) and travel along the antenna surface to the edges of the cone and disc. In designing the discone, the dimensions of the antenna are carefully computed so as to make the impedance at its edges similar to that of free space. Naturally the discone radiates a signal because there is a maximum transfer of energy when impedances are matched.

The discone antenna described in this chapter is made of wirescreen mesh. This material is purposely used to minimize the effect of wind to the antenna. The thin metal strips used to clamp the two overlapping edges of the cone is for mechanical reasons only- the RF waves travel down to the cone edge and not around it, so an electrical connection is not important.

This unit has an operational frequency bandwidth of 140 MHz up to 1.4 Gigahertz although best results can be obtained if its use is limited up to 1 Gigahertz only. SWR is measured to be less than 2:1 over the entire bandwidth. Power gain is 1 dB (unity gain) and its installation is fixed. The discone antenna does not need tuning after construction. It is also popular for use in automatic scanning wideband monitors.

SCALED DOWN APPEARANCE

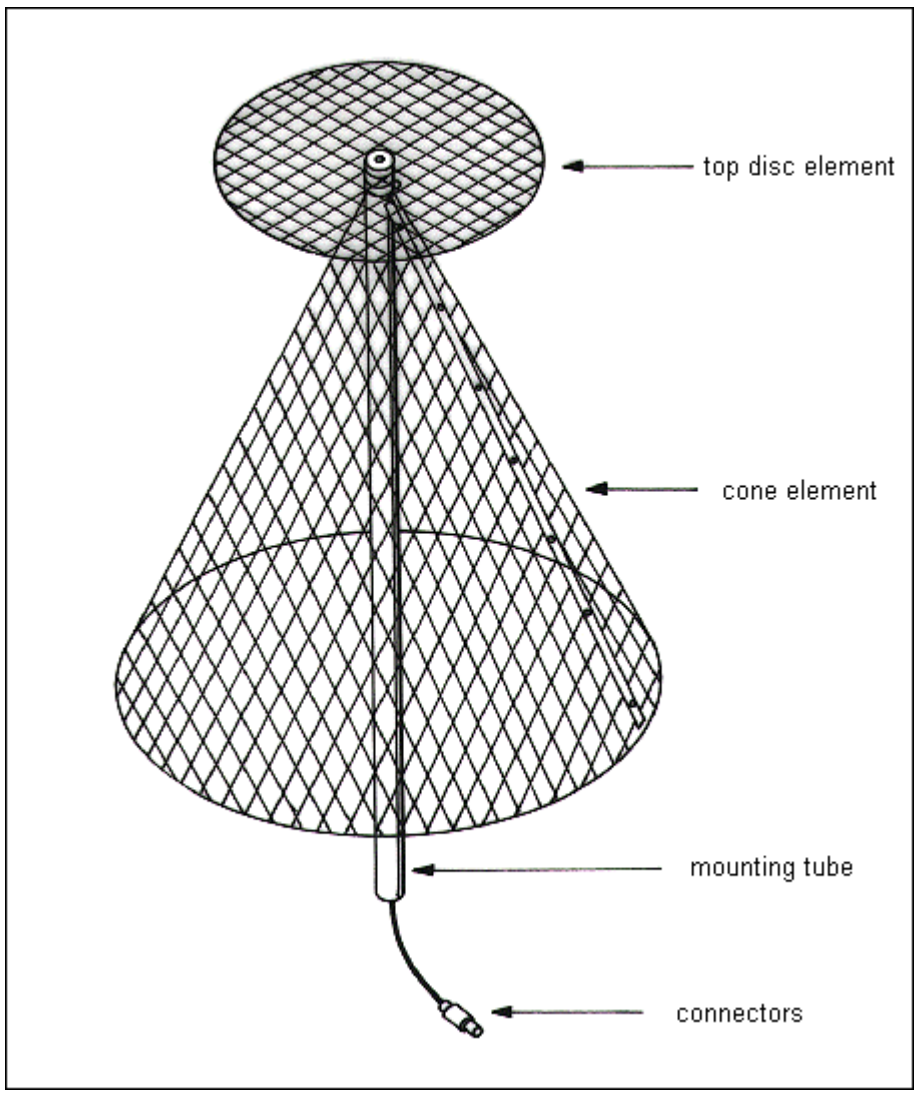


Figure 8.1

Materials List

Quantity	Specification/Description	Dimensions
1	Aluminum or GI wire screenmesh medium gauge - enough to support itself without reinforcement	
1	Aluminum tube 1" id*	25" long
1	PL-259 VHF male connector	
1	PL-258 VHF straight connector	
1	Coaxial cable RG-58/U	30" long
2	Eye terminals 1/8"	
1	Washer - aluminum (customized dimensions see text)	
1	Plastic bushing (customized dimensions see text)	
1	Stove bolt - brass or GI	1/8" x 1-3/4"+
1	Hex nut 1/8" id*	
1	Hose clamp - stainless steel	1-1/2" diameter
1	Metal plate 1/8" thick	3" x 6"
2	Aluminum strip gauge 14 or 16	1/2" x 22"
4	U-bolt with hex nuts and lock-washers	

*id- inside diameter

Construction

First, prepare the customized aluminum washer to be used as a disc holder. Machine it from a thick aluminum plate or rod following the dimensions shown in Figure 8.2.

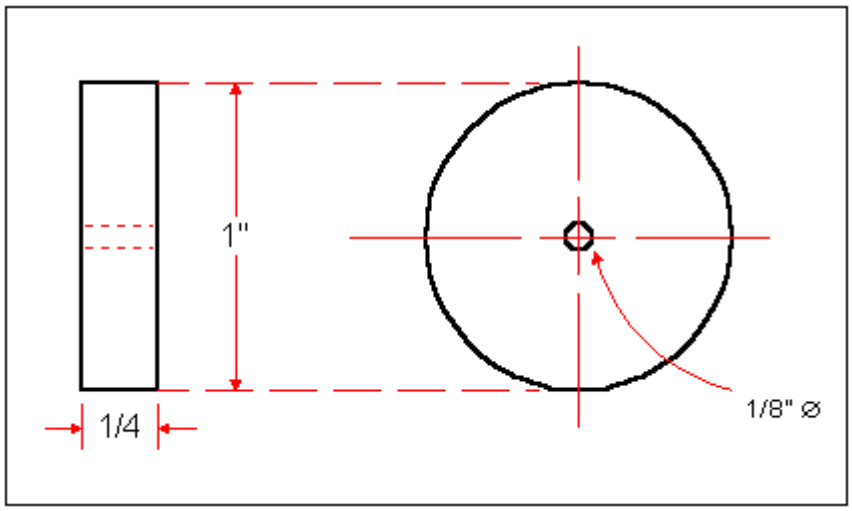


Figure 8.2 Customized aluminum washer dimensions.

Next, prepare the plastic bushing from a small piece of engineering plastic rod with the required diameter. Machine it according to the dimensions shown in Figure 8.3.

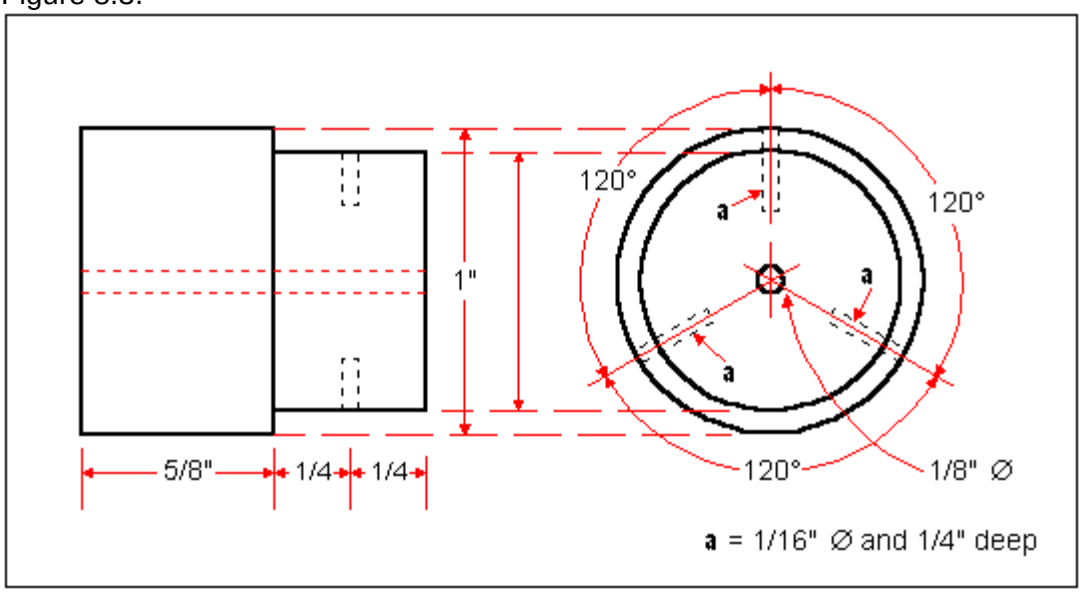


Figure 8.3 Plastic bushing dimensions.

Next, prepare the aluminum mounting tube. The tube must be 1" in diameter and 25 inches long. Drill three holes (1/8" diameter) around one end of the tube with the holes equally spaced between each other (see Figure 8.4).

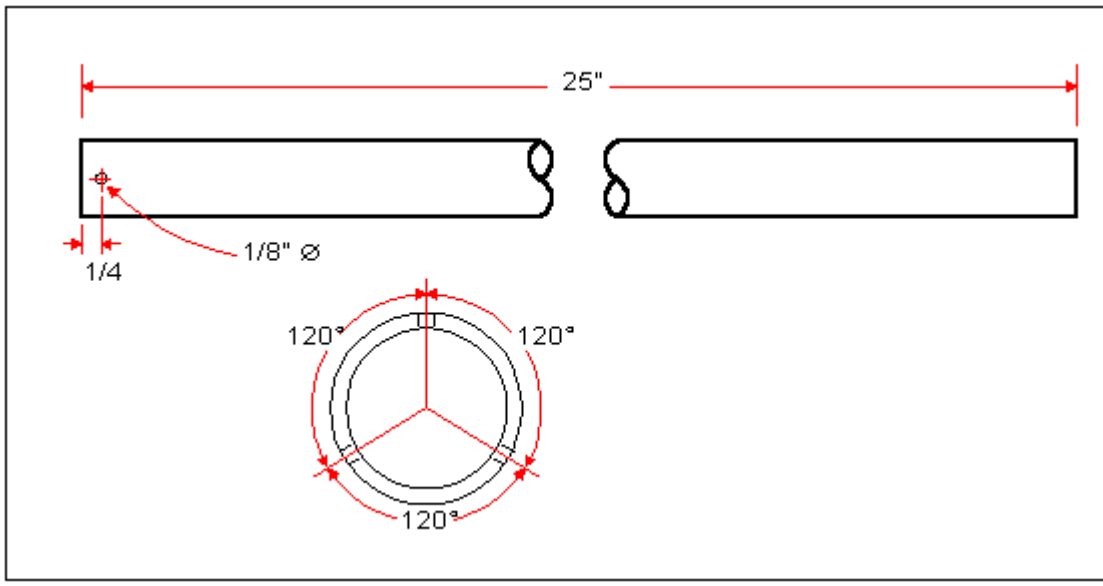


Figure 8.4 Drilling holes at one end of the mounting tube.

Drill a single hole (1/8" diameter) at the same end but slightly lower than the first three holes (see Figure 8.5). This hole will accomodate the screw to hold the coaxial braid inside the tube as described later in the final steps.

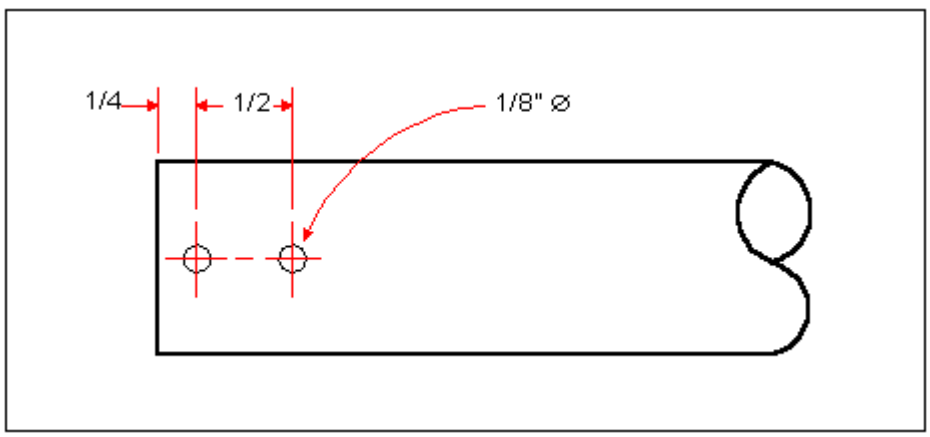


Figure 8.5 Drilling the hole which accomodates the coaxial braid lockscrew.

Preparing the disc and cone

Cut the disc element from the aluminum screen mesh using a suitable tinsnip (see Figure 8.6).

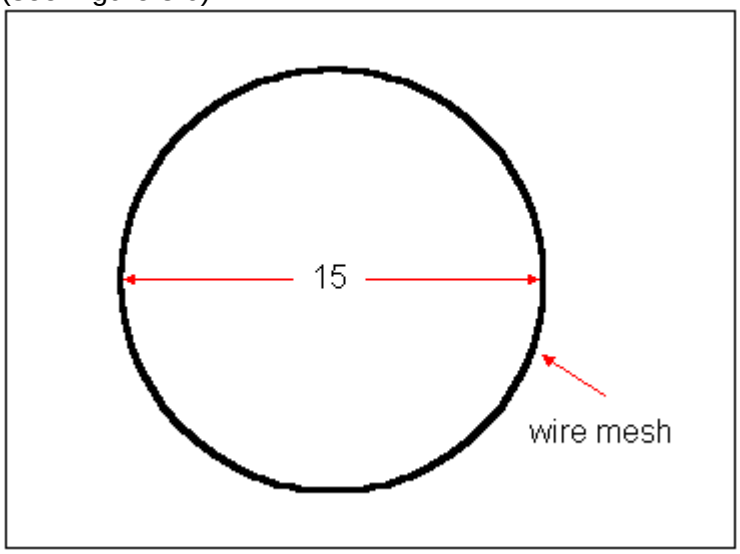


Figure 8.6 Disc element dimension.

Next, prepare the cone element from a similar material. Follow the dimensions shown in Figure 8.7.

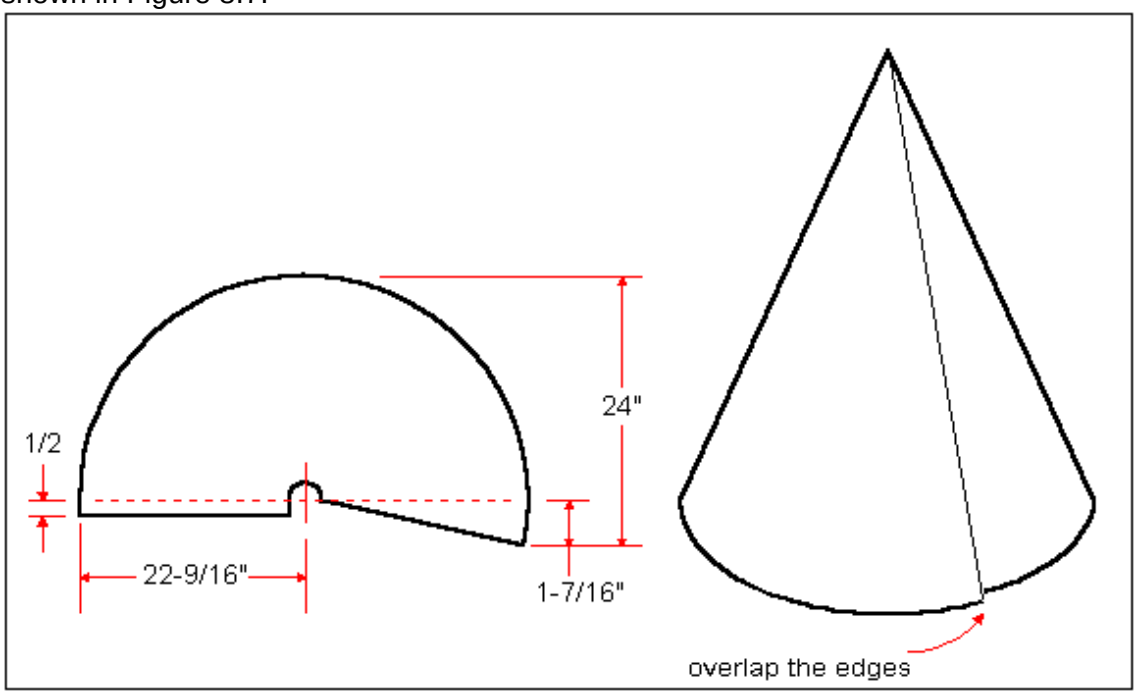


Figure 8.7 Cone element dimensions.

Prepare the aluminum strips according to the dimensions shown in Figure 8.8. These two strips will be used to clamp the two overlapping edges of the cone permanently.

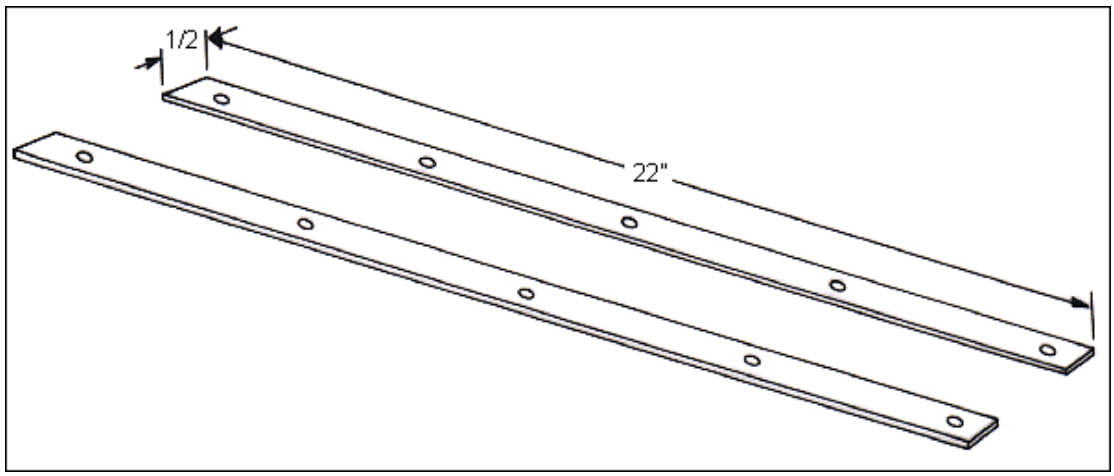


Figure 8.8 Preparing the clamping strips.

Place one strip along the overlapping edges under the cone and place the other strip over the overlap outside the cone. Align the holes in both strips and rivet the two pieces together. The rivet must pierce through the two overlapping edges (see Figure 8.9). The riveted strips must sandwich the screen mesh and hold the cone form rigidly.

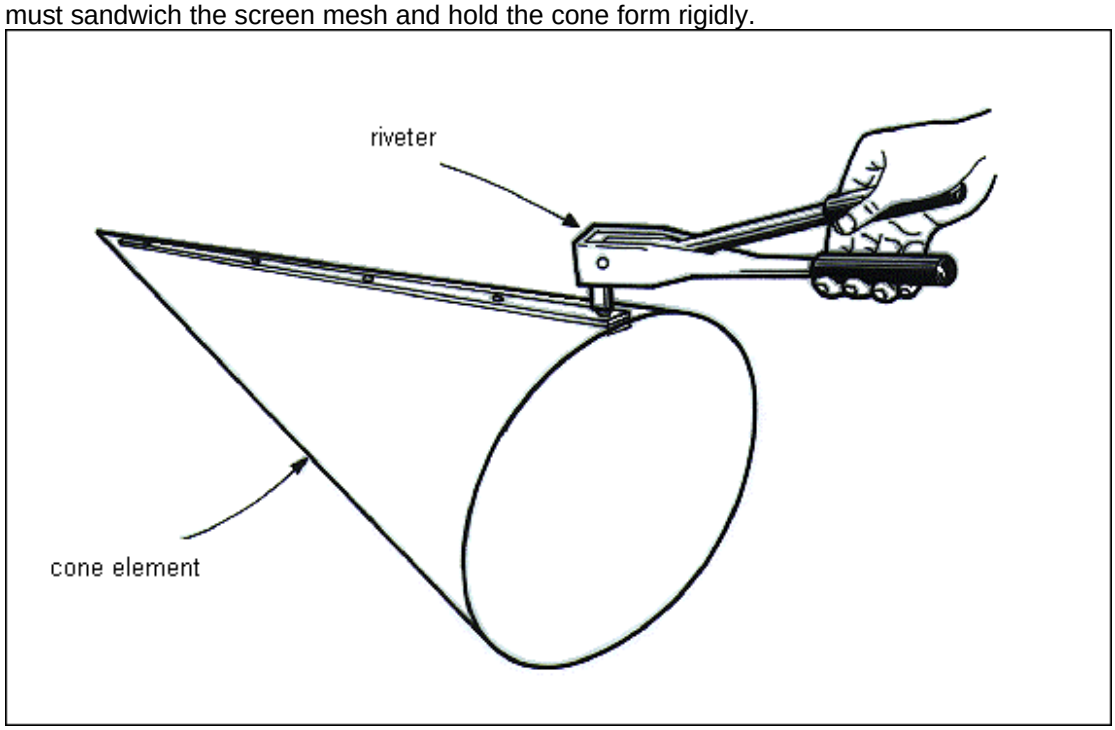


Figure 8.9 Riveting the clamping strips.

By using a tinsnip, make crosscuts on the apex of the cone to make a hole large enough for the mounting tube to go through. Follow the illustration in Figure 8.10 carefully . The cuts must result to an opening equal to the diameter of the tube - around 1 inch.

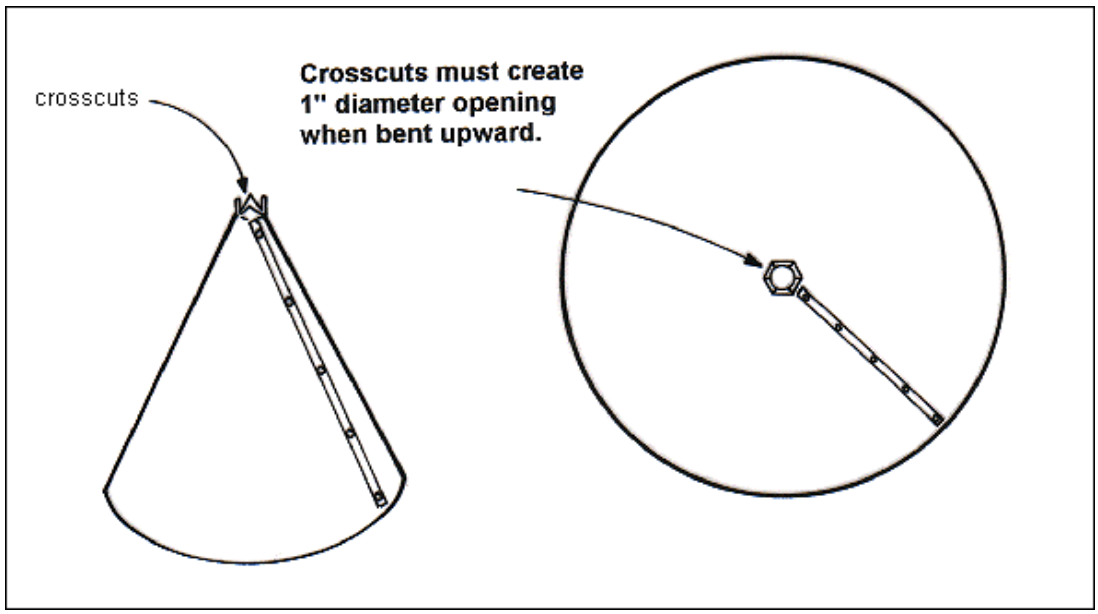


Figure 8.10 Making crosscuts at the apex of the cone.

Assemble the top disc elements following the illustrated steps in Figure 8.11.

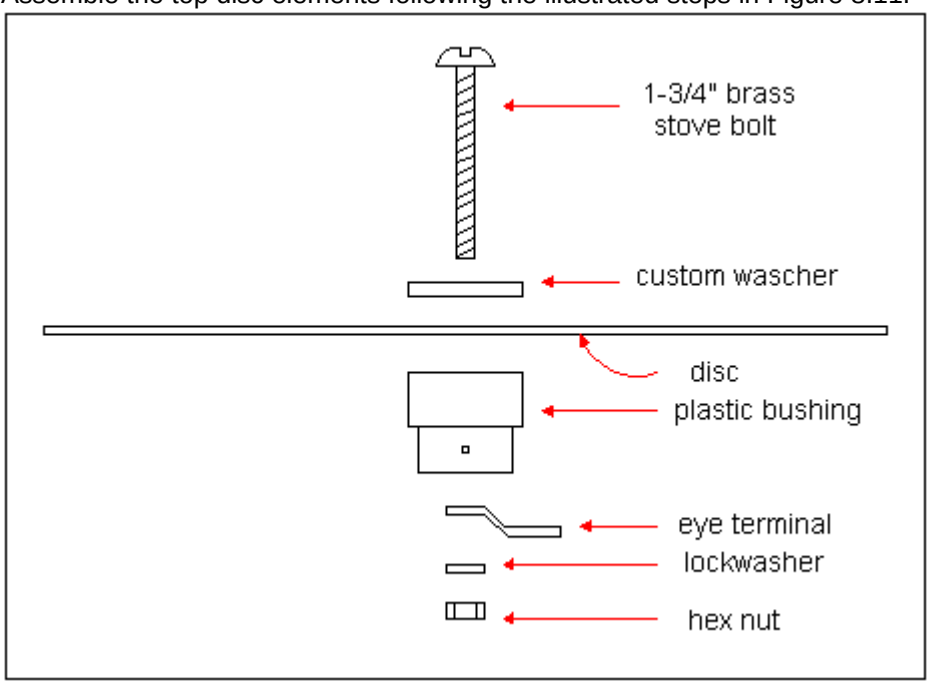


Figure 8.11 Assembling the top disc hardware.

Cut a 30 inches long RG-58/U coaxial cable and solder the inner conductor at one of its ends into the eye terminal held by the bolt below the plastic spacer. Attach a vinyl insulated eye terminal into its braid (see Figure 8.12).

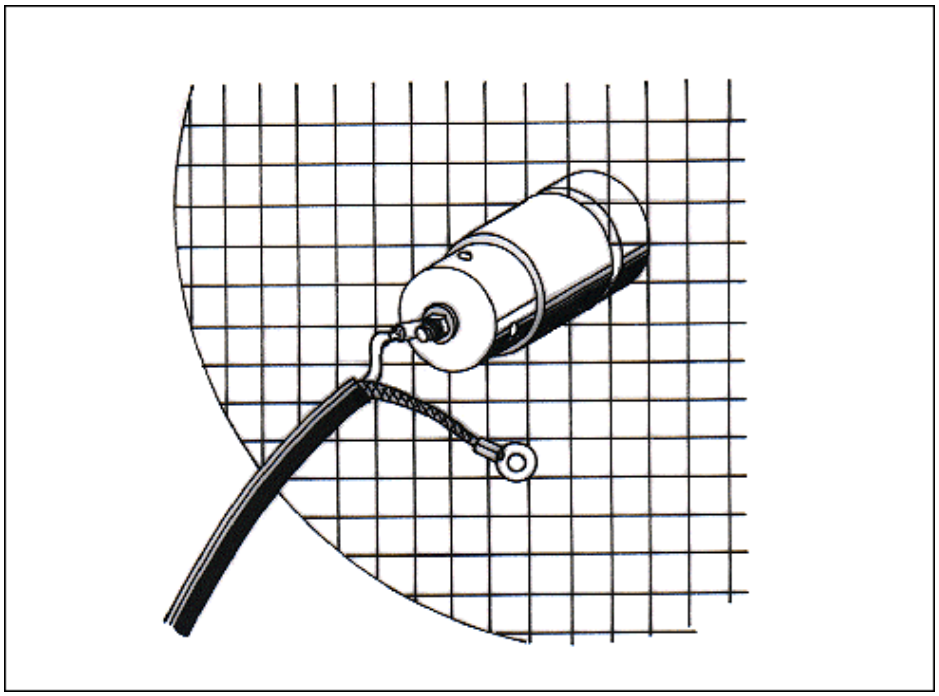


Figure 8.12 Connecting the coaxial cable to the top disc element.

Insert the free end of the coaxial cable into the mounting tube starting from the tube's end with sideholes (see Figure 8.13).

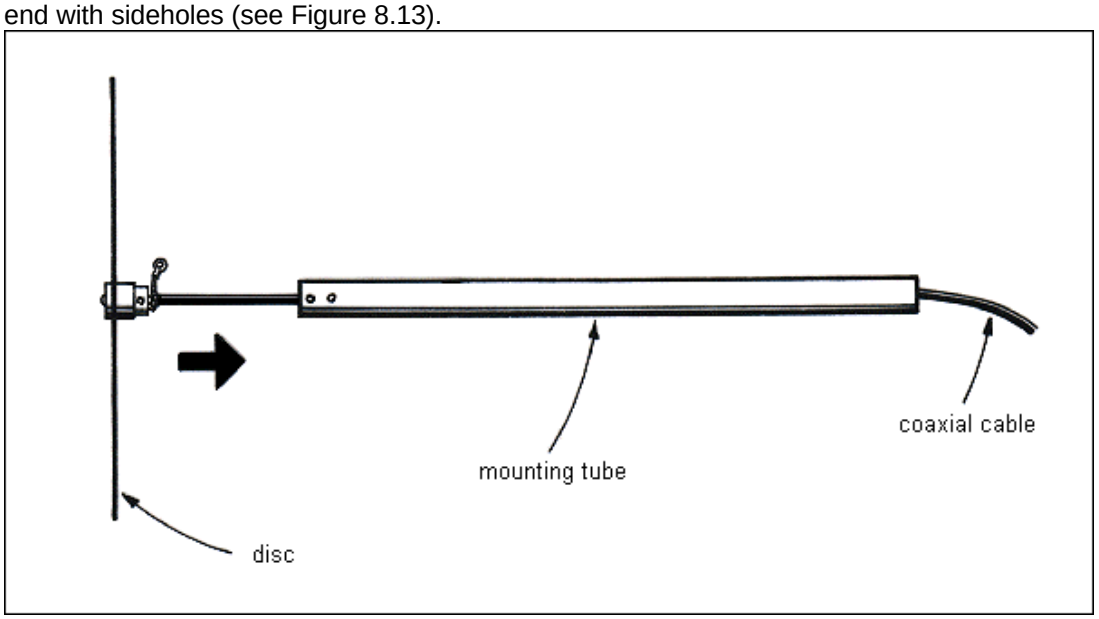


Figure 8.13 Inserting the coaxial cable into the mounting tube.

Insert the plastic spacer/bushing holding the top disc element inside the tube and align its hole to the tube's sideholes. Fix the bushing to the tube per-manently using metal screws (see Figure 8.14).

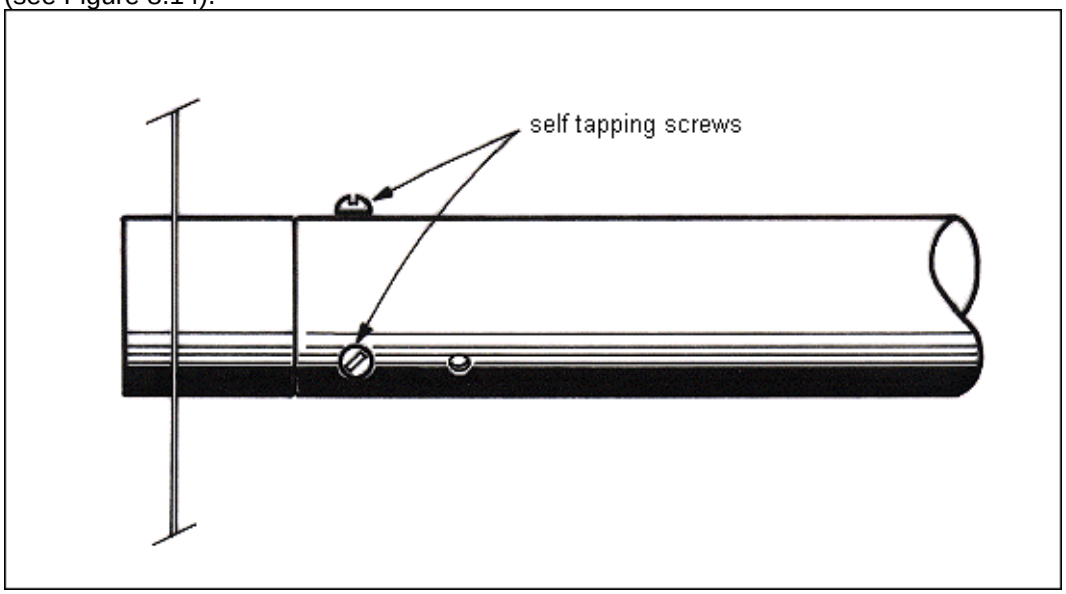


Figure 8.14 Securing the top disc element to the tube.

Align the lower hole to the eye terminal of the braid inside the tube. If it is not aligned yet, insert a slender stick inside the tube and remotely move the eye terminal until you can see it through the hole outside. Insert a metal screw into the hole and turn it until it catches the eye terminal inside. Tighten the screw to hold the terminal firmly (see Figure 8.15). This procedure is the main reason why you should use an eye terminal with a 1/16" diameter eye.

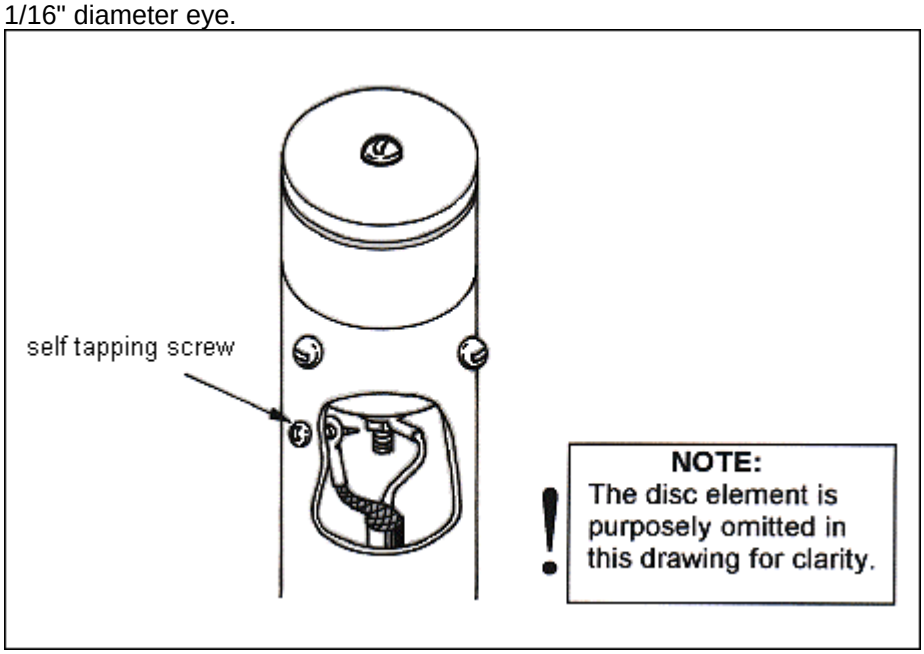


Figure 8.15 Securing the braid inside the mounting tube.

Solder a PL-259 to the free end of the coax cable and connect a straight connector (PL-258) into it prior to the final installation of the antenna (see Figure 8.16).

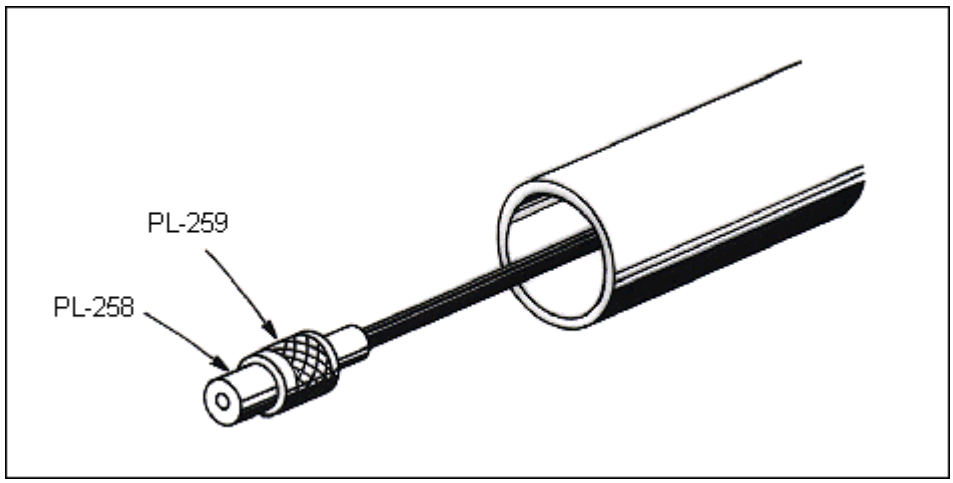


Figure 8.16 Connecting the PL-259 and PL-258 to the coaxial cable.

Finally, insert the PL-259, coaxial cable and the mounting tube into the cone starting from the top until the apex of the cone reaches just a tiny fraction of an inch below the plastic bushing holding the top disc element (see Figure 8.17).

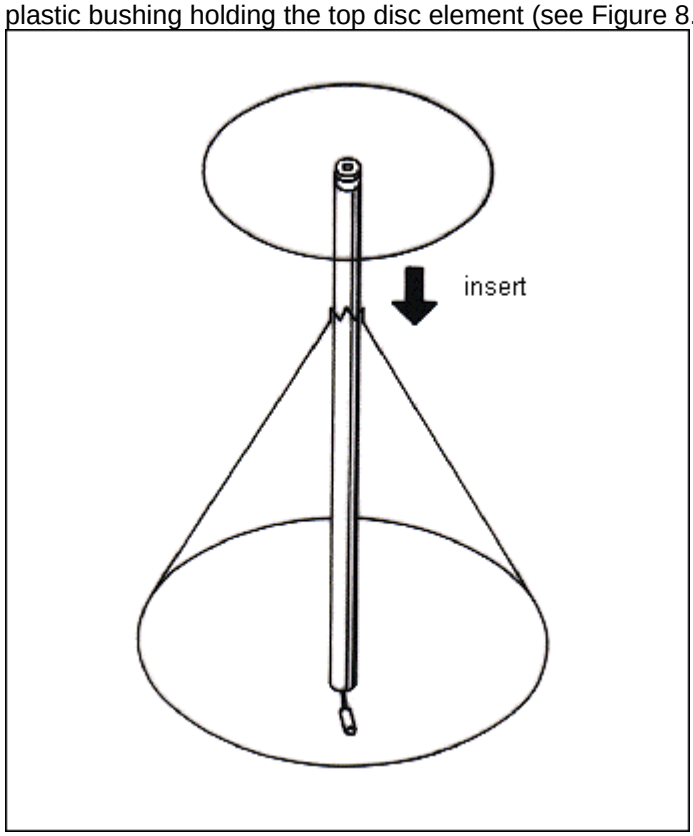


Figure 8.17 Inserting the mounting tube into the cone element.

Attach the stainless clamp around the upturned portion of the wiremesh just under the disc. Tighten the clamp to hold the cone in place. Trim the excess wiremesh protruding above the edge of the tube clamp (see Figure 8.18).

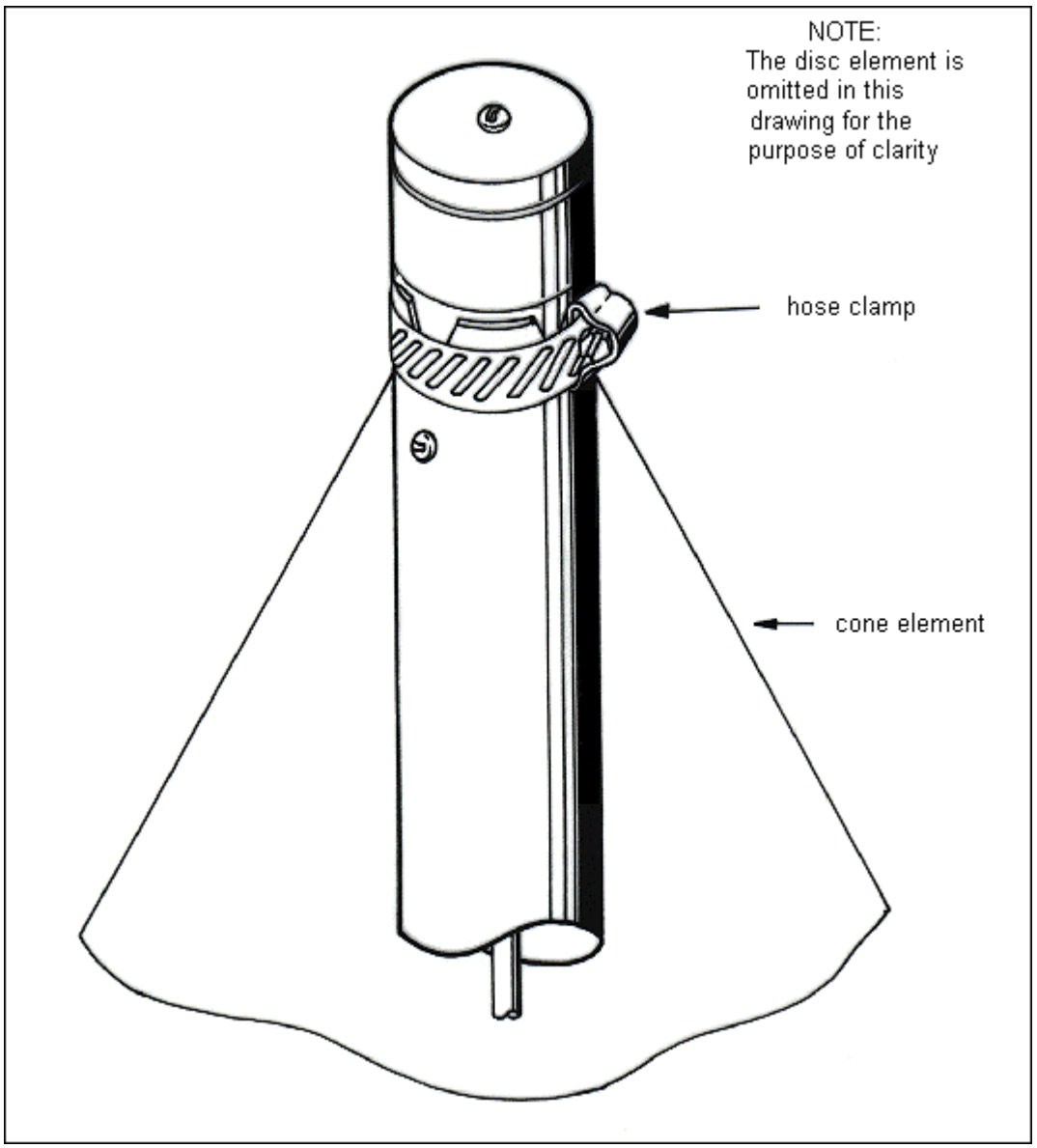


Figure 8.18 Clamping the apex of the cone element to the mounting tube.

Installation of CD-2W

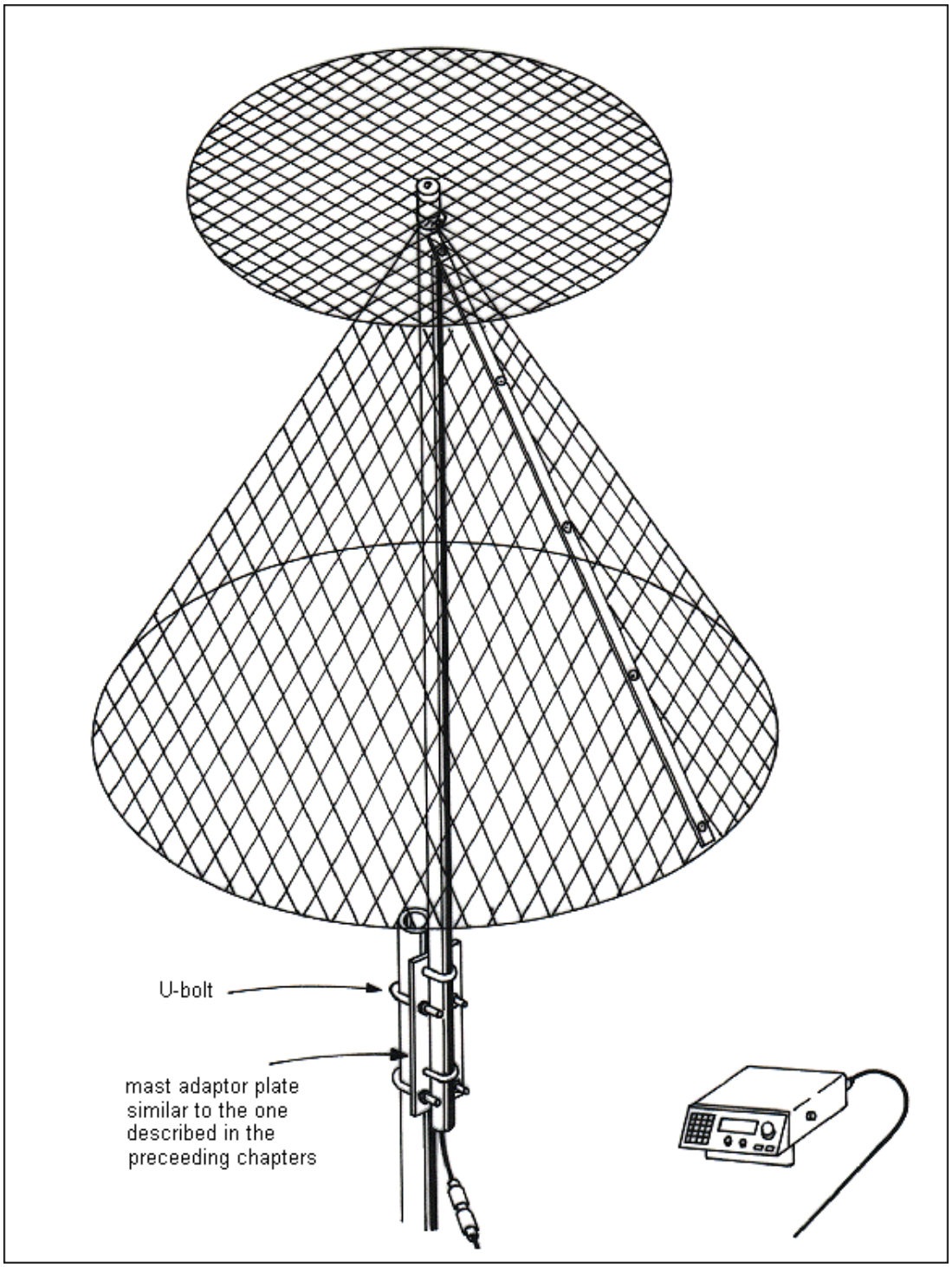
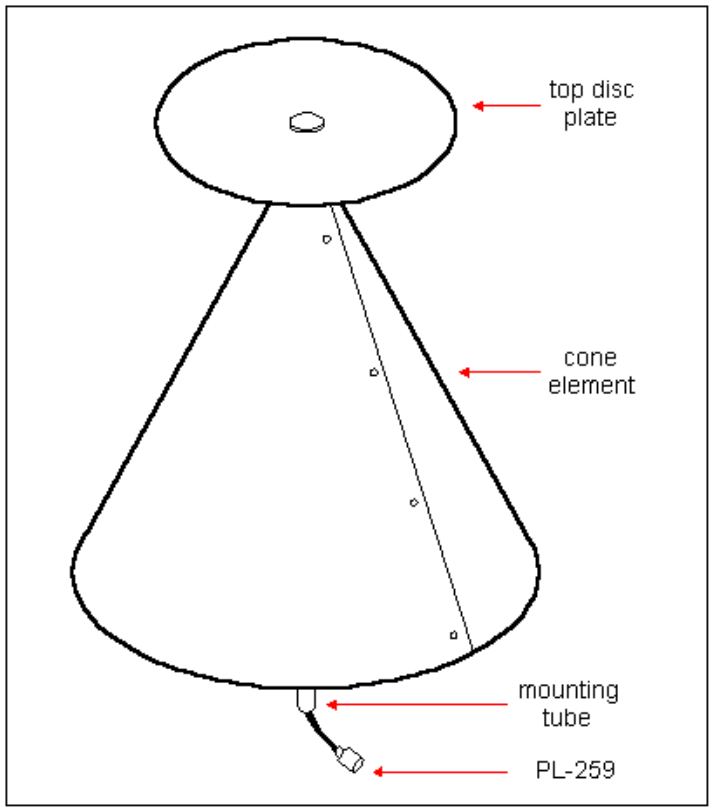


Figure 8.19 Mounting the CD-2W to the mast.

9 DISCONE ANTENNA

Model CD-2P



The discone antenna model CD-2P described in this chapter is functionally similar in most respects to the discone antenna in chapter 8. The only and obvious difference between the two models is the utilization of a metal plate for the disc and cone elements of CD-2P (P for plate).

The choice of using a metal plate becomes evident when the antenna is intended to be installed in areas with less than excellent weather conditions - meaning if your area is regularly visited by heavy rainfall or strong winds, then you must opt to construct and install this more robust model than the wire-screen version. Metal plate is more durable than a wire-screen. The only trade-off is the total cost of the antenna because metal plate is more expensive. In most occasions, a GI metal plate is satisfactory but you can also use a more expensive aluminum plate if you desire so. Aluminum is less susceptible to corrosion so it is highly recommended if you plan to use the antenna near seashores or in places where there is a high level of salt present in the surrounding moisture.

Paint has a negligible effect on the RF signal so if you decide to paint the antenna to make it look attractive, just do it - but don't paint the aluminum mounting tube under the cone element to ensure adequate grounding connection of the antenna to the mast or tower. This is a precautionary measure to avoid a lightning striking your antenna and possibly causing damage to your transceiver or to you.

Materials needed

The materials necessary for the antenna Model CD-2P are the same with those needed for CD-2W except for the wire mesh for the cone and disc which is replaced with a thin metal sheet in CD-2P. Also the two thin metal strips are not needed. However you have to retain the blind rivets for the same purpose.

Construction

In constructing most of the parts of CD-2P, follow the instructions for the Model CD-2W except those for the disc and cone elements. Cut the cone and disc elements out of the thin metal sheet following the dimensions shown in Figures 9.2 and 9.3.

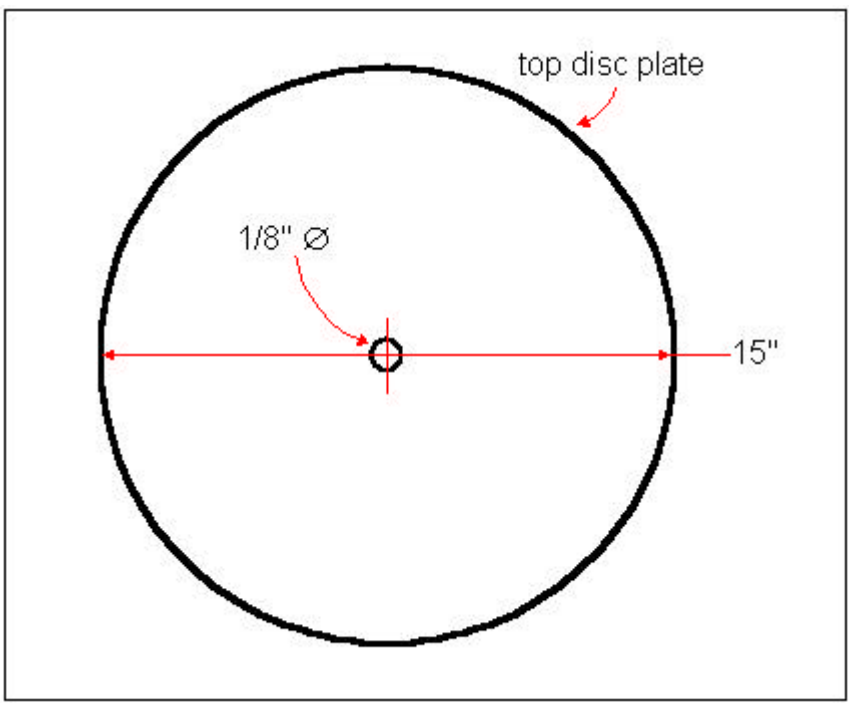


Figure 9.2 Disc element dimension.

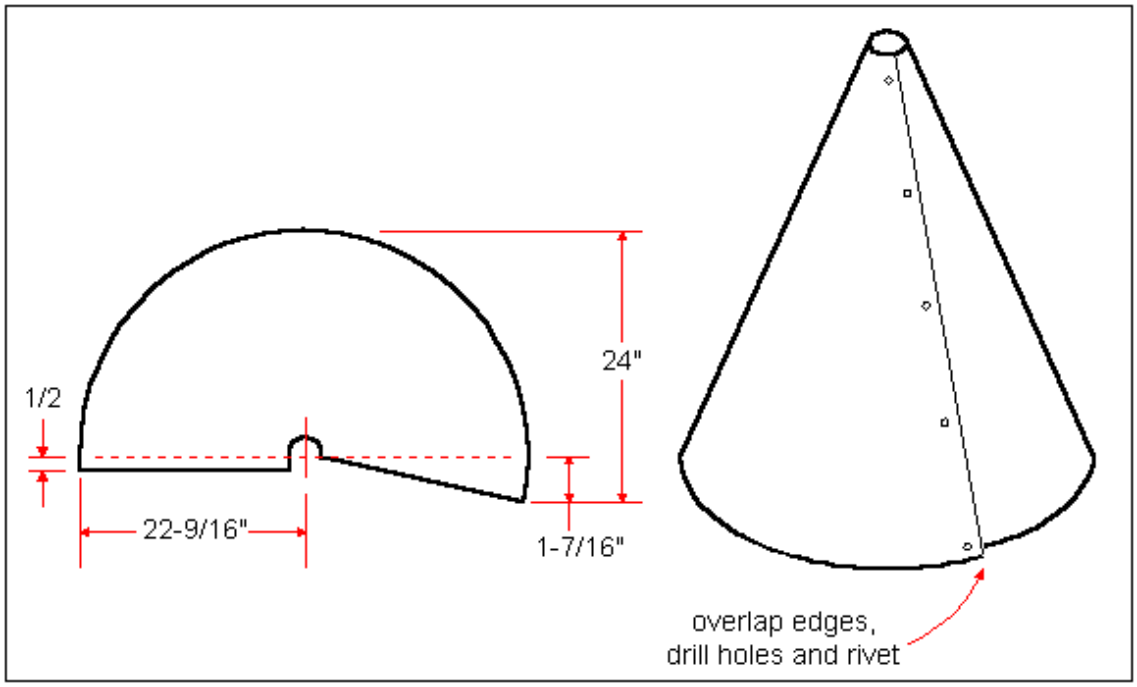


Figure 9.3 Cone plate dimensions.

Unlike the wire mesh cone of CD-2W, the cone for CD-2P must have a metal sleeve soldered to the opening at the apex. Cut a metal sheet and shape it to a form of a ring as shown in Figure 9.4.

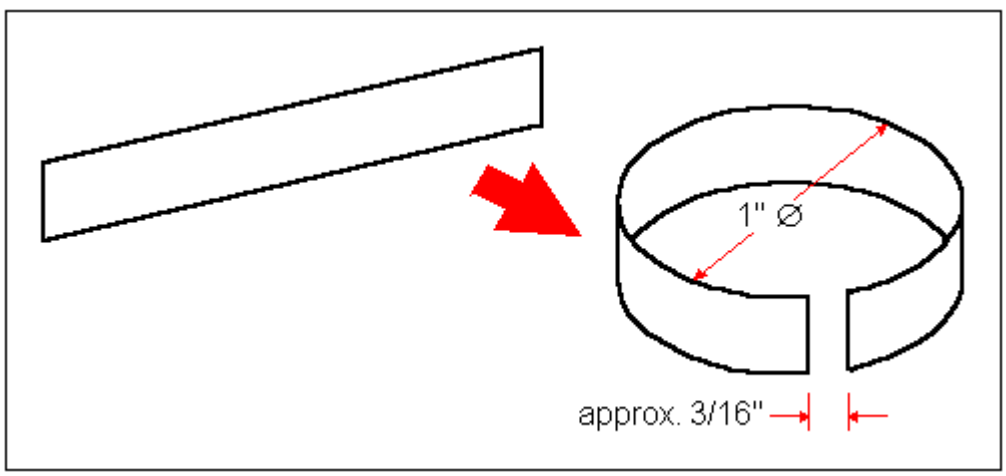


Figure 9.4 Preparing the sleeve.

Solder this sleeve to the rim of the apex opening leaving a small gap between their ends (see Figure 9.5). If you are using an aluminum plate for the cone and sleeve, you need to electrically weld the two pieces together using a special technique for welding aluminum with protective gas.

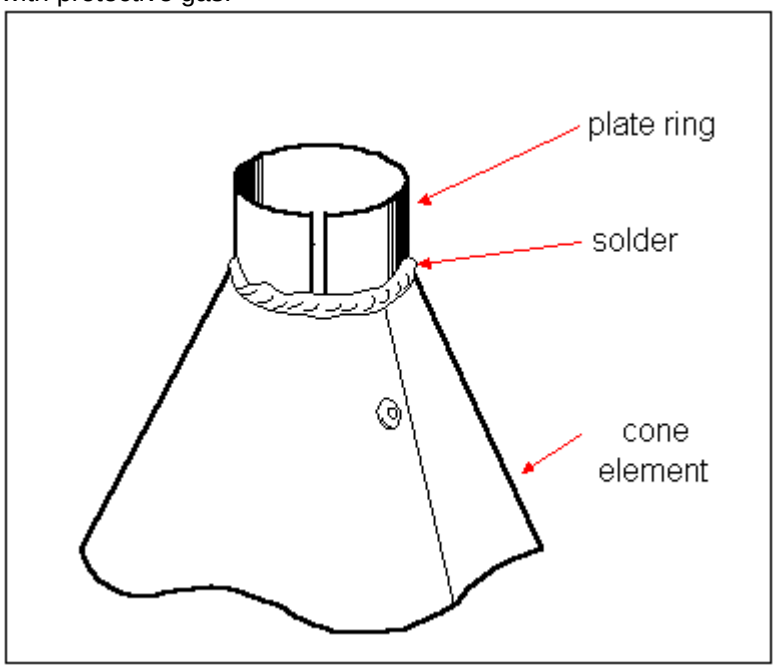


Figure 9.5 Soldering or welding the sleeve to the apex opening of the cone.

The assembly of CD-2P is similar to the steps for assembling the CD-2W (see Figure 9.6).

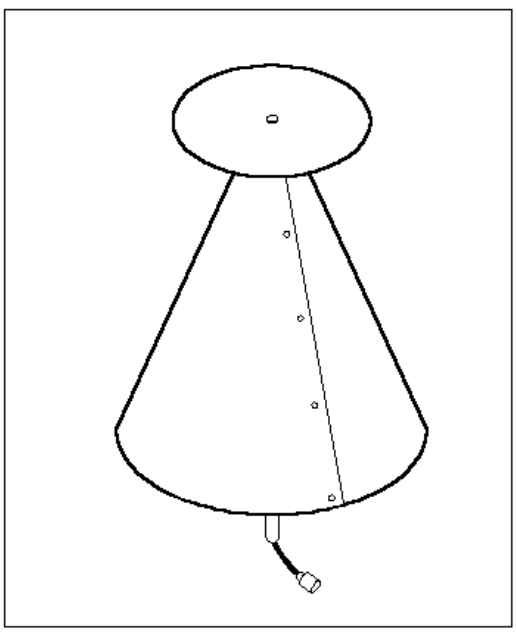


Figure 9.7 Assembled CD-2P.

10 DISCONE ANTENNA

Model CD-2T

In mobile operations the problems related to antenna installation are much greater than those encountered in fixed stations. The problems are worse particularly for a one-man-mobile-station which travels on foot. Those seemingly small items like portable transceiver, spare batteries, coaxial cable, myriad of wires, solar panels, charging box, log books, scanning monitor, etc. could easily total up to more than 20 kilos of deadweight if cramped together inside a single backpack. Add to it the supply of food and few personal belongings and it will surely feel like a nightmare when travelling across a rugged terrain.

The over-all bulk of the load is another problem. Just imagine travelling while lugging a full-size metal plate discone at your back! Because of this, the tendency of mobile operators is to bring only the most important piece of equipment and is usually a portable and lightweight version to trim down the total weight and bulk of the load.

The antenna model described in this chapter is specially designed to satisfy the need for a lightweight and transportable discone antenna. The cone and disc elements are replaced with retractable telescopic rods so that the antenna can be collapsed into a small unit and conveniently stored inside a backpack. The actual length of a packed discone is merely 8 inches! When the telescopic rods are extended to their maximum length and set in the proper angle, they approximate the function of a full disc and cone elements. Theoritically the more elements used the better. Experience showed however that three elements for each disc and cone function are enough on most occasions.

This portable version of a discone antenna has the same electrical characteristics with the two full-sized models described in chapters 8 and 9. The only difference is in the mechanical construction. The chrome plated telescopic rods are quite expensive so the total cost of this antenna is higher than the two preceeding models.

If this antenna will be used solely for mobile operations, then the U-bolts intended for mounting may be discarded. Instead, the antenna can be tied with a thin nylon rope on its mounting tube near the feedpoint and hung under a post or branch of a tree. Never use a metallic wire to hang the antenna because it will distort the radiation pattern of the signal or short out the disc and cone elements.

Scaled down appearance

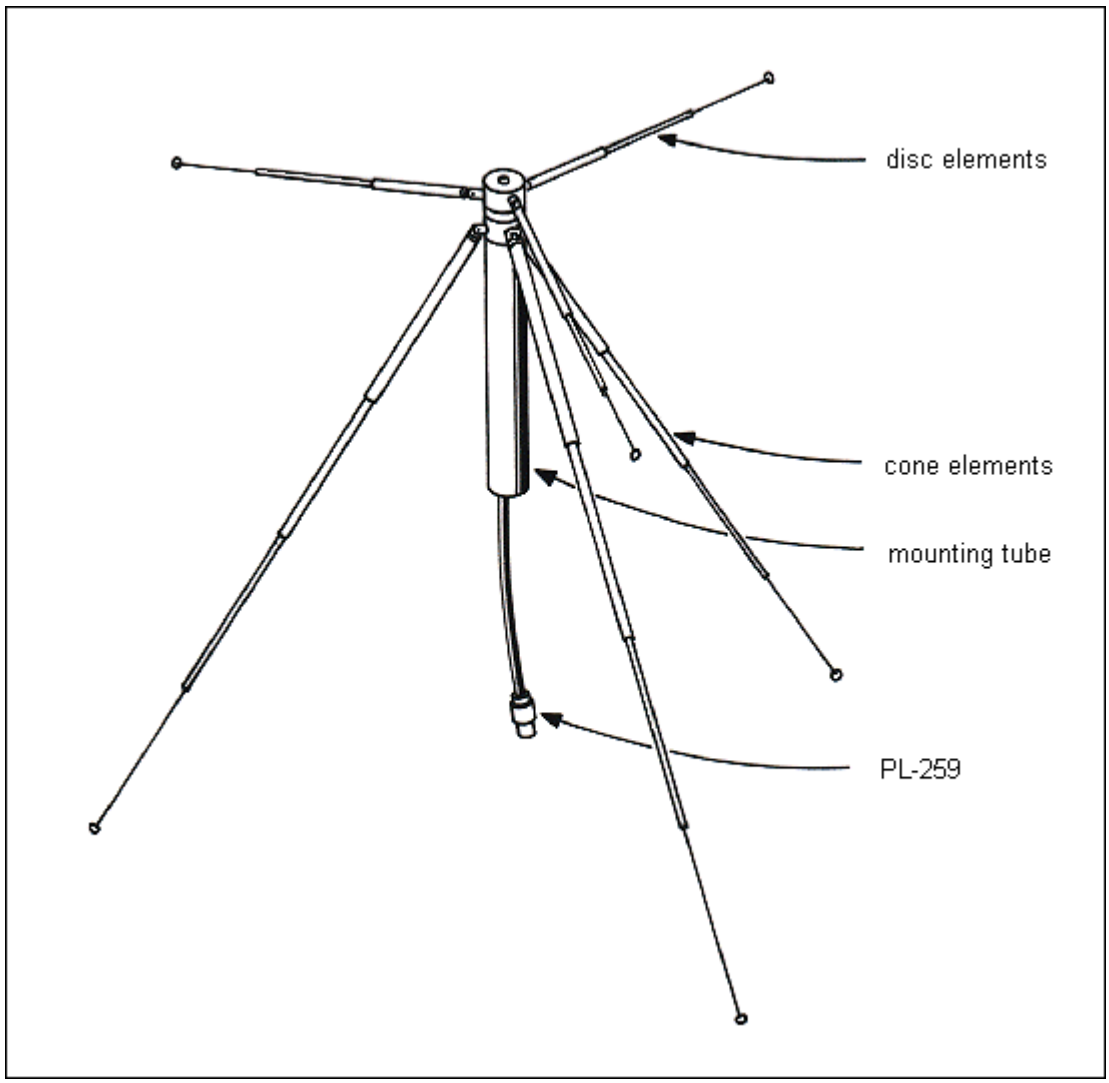


Figure 10.1

Materials List

Quantity	Specification/Description	Dimensions
3	Telescopic antennas - with swivelling threaded base	7" fully extended 3-5" retracted
3	Telescopic antennas - with swivelling threaded base	22" fully extended 5-6" retracted
1	Aluminum disc base mount see text for exact dimensions	
1	Plastic spacer - see text for customized dimensions	
1	Aluminum cone base mount see text for exact dimensions	
1	PL-259 VHF male connector	
1	PL-258 VHF straight connector	
1	Aluminum plate 1/8" thick	3" x 6"
2	Eye terminals - no insulation	
4	U-bolts with accompanying hex nuts and lockwashers	
1	Stove bolt - brass or GI	1/8" x 2"
6	Self tapping metal screws	1/8" x 3/8"
1	Self tapping metal screw	1/8" x 3/4"
1	Hex nut - brass or GI	1/8" id*

*id- inside diameter

Construction

First, prepare the top disc elements base mount by machining an aluminum rod to the necessary dimensions. Follow the dimensions shown in the illustration (Figure 10.2). The size of the holes (holes marked with **a**) and their thread gauge must conform to the dimensions of the short telescopic antenna you intend to use.

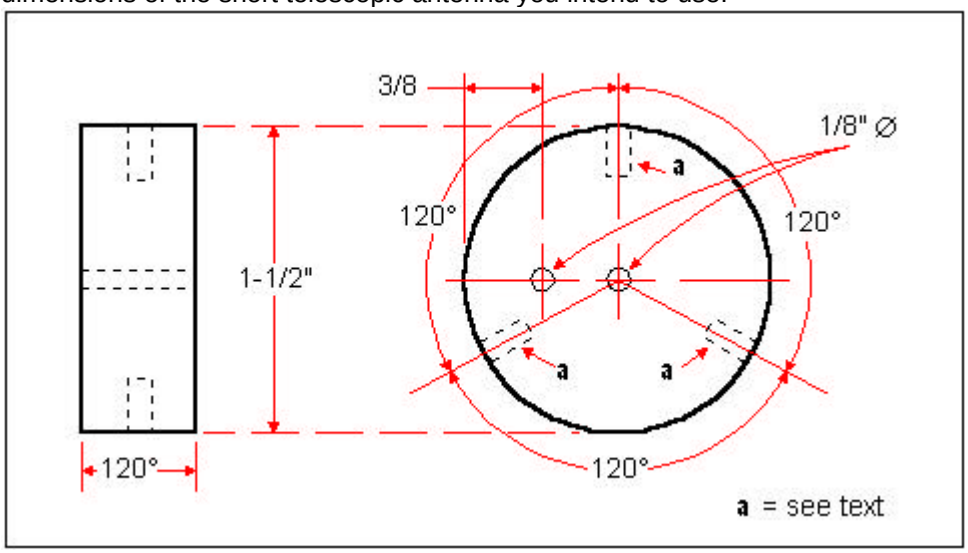


Figure 10.2 Disc base mount dimensions.

Second, prepare the plastic spacer/bushing from a piece of an engineering plastic. Machine it to the form and dimensions shown in Figure 10.3. This plastic spacer insulates the disc base mount from the cone base mount.

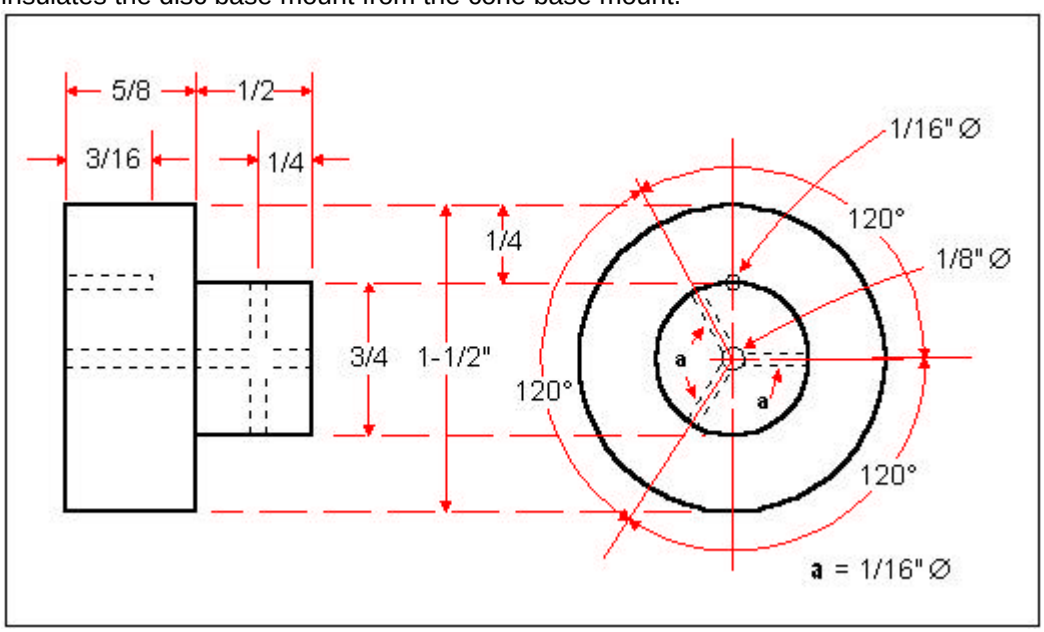


Figure 10.3 Plastic spacer dimensions.

Next step is to prepare the cone elements base mount from an aluminum rod with the necessary size. Machine it to the form and dimensions shown in the following illustration (Figure 10.4). The size, thread gauge, and deepness of the holes at the side (holes marked with **a**) must conform to the base dimensions of the particular type of telescopic antenna intended for the cone elements.

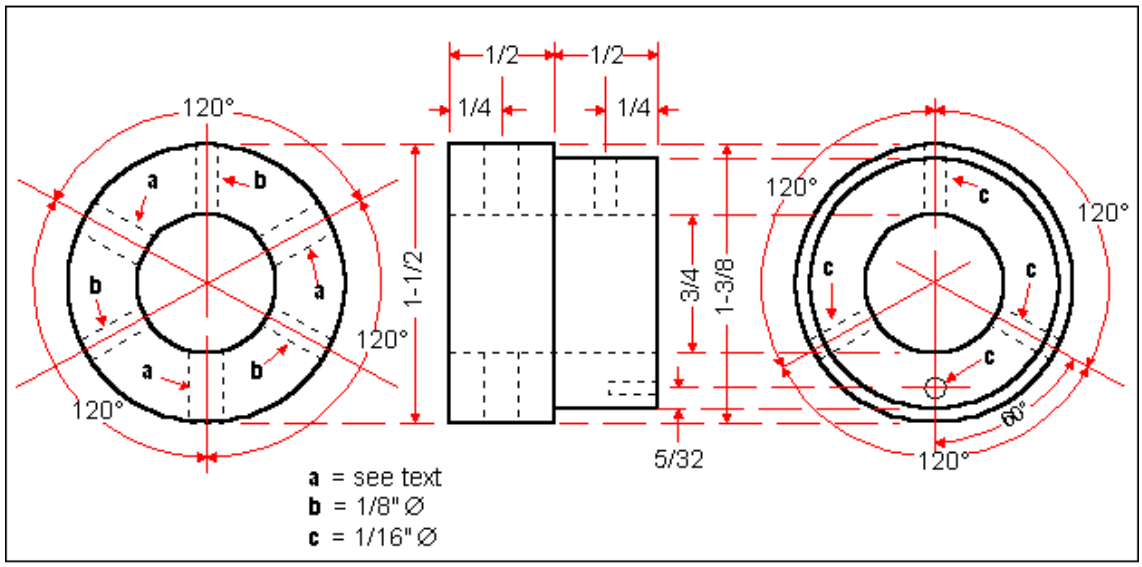


Figure 10.4 Cone elements base mount dimensions.

Assemble the disc base mount, the plastic spacer and the cone base mount together following the arrangement shown in Figure 10.5. Secure the assembly with self tapping screws to the appropriate holes as illustrated.

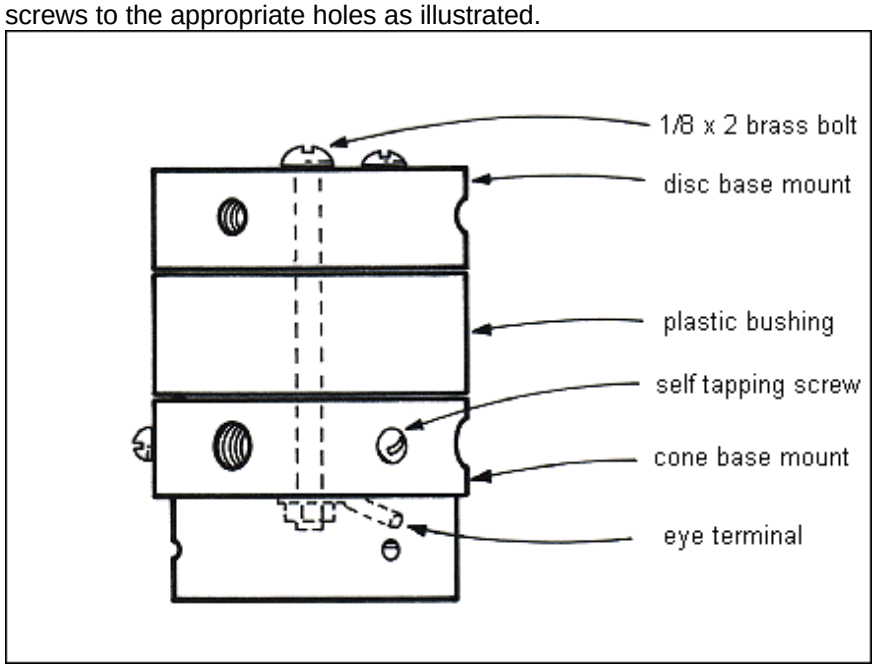


Figure 10.5 Assembly of the elements' base mounts.

Attach the remaining eye terminal to the lone hole at the rim of the cone base mount (see Figure 10.6).

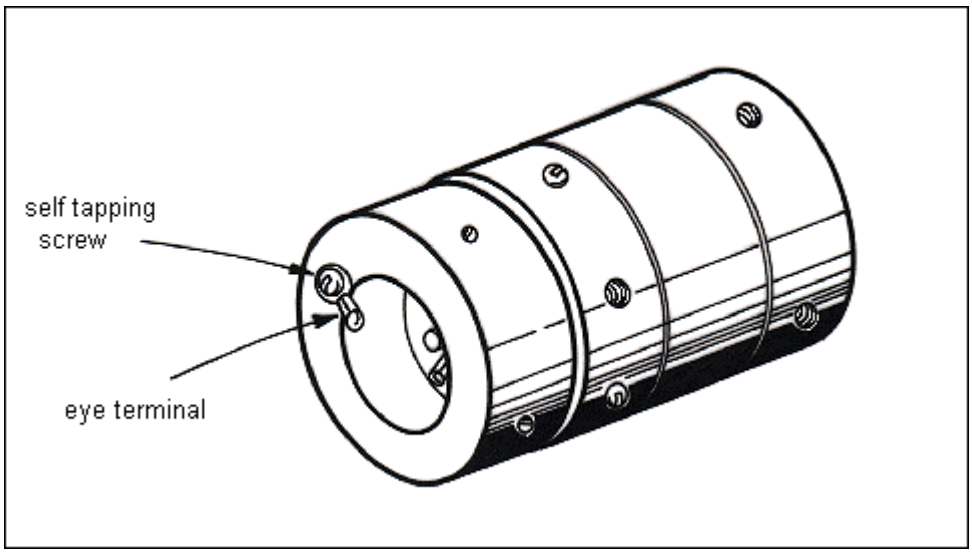


Figure 10.6 Eye terminal attached to the rim of the cone base mount.

Solder one end of the coax cable to the two terminals at the base mount assembly. The inner conductor must be soldered to the center terminal and the braid must be soldered to the eye terminal at the rim (see Figure 10.7). **Never interchange the connection**.

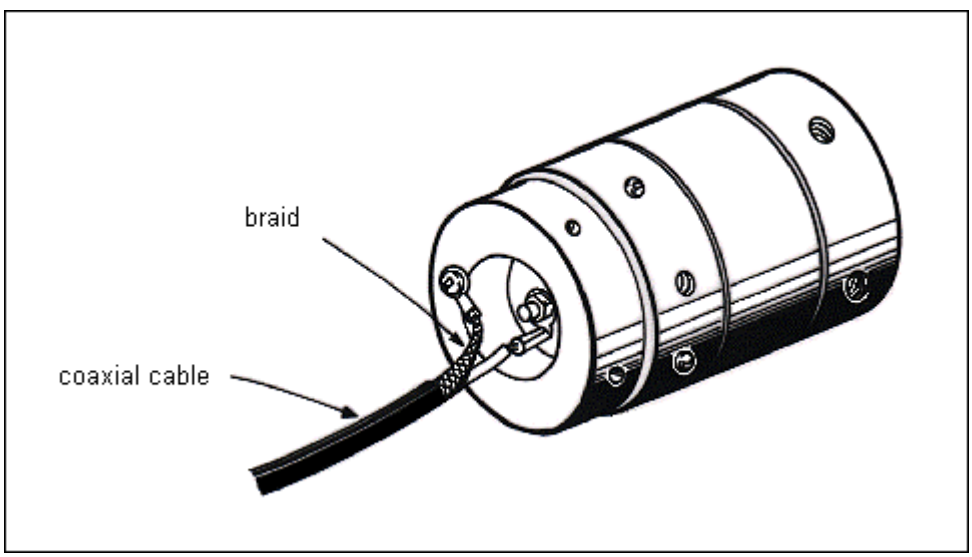


Figure 10.7 Connecting the coaxial cable to the terminals.

Next step is to prepare the aluminum mounting tube by cutting it to a length of 6 inches. Drill three holes (1/8" diameter) at one end (see Figure 10.8). The holes must be equally spaced from each other.

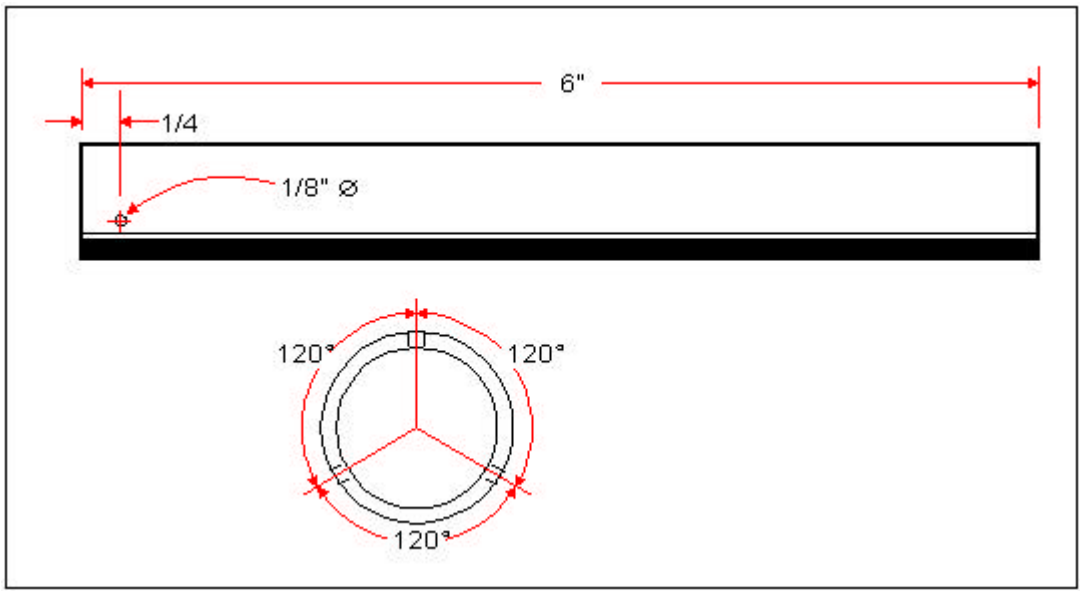


Figure 10.8 Preparing the mounting tube.

Insert the free end of the coax cable inside the aluminum tube starting at the end with sideholes. Insert the aluminum base mount assembly into the tube and align the holes at the sides. Place screws through the holes to permanently attach the base mount assembly into the tube (see Figure 10.9).

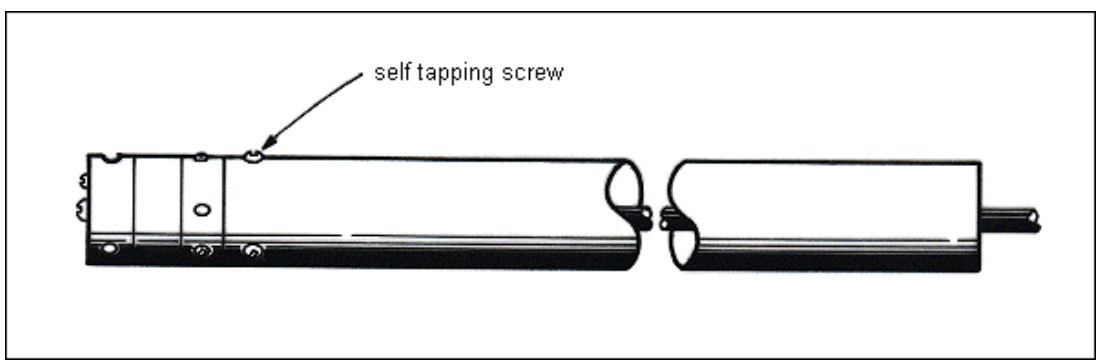


Figure 10.9 Fixing the base mount assembly into the tube.

Solder the PL-259 to the free end of the coaxial cable and attach a straight connector (PL-258) into it (see Figure 10.10).

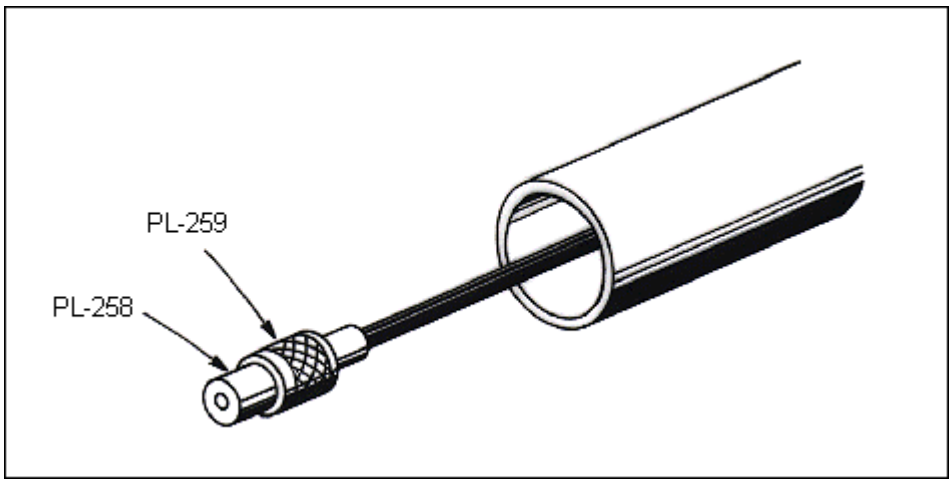


Figure 10.10 Connecting the PL-259 and PL-258.

Attach the three short telescopic antennas into the disc base mount (see Figure 10.11).

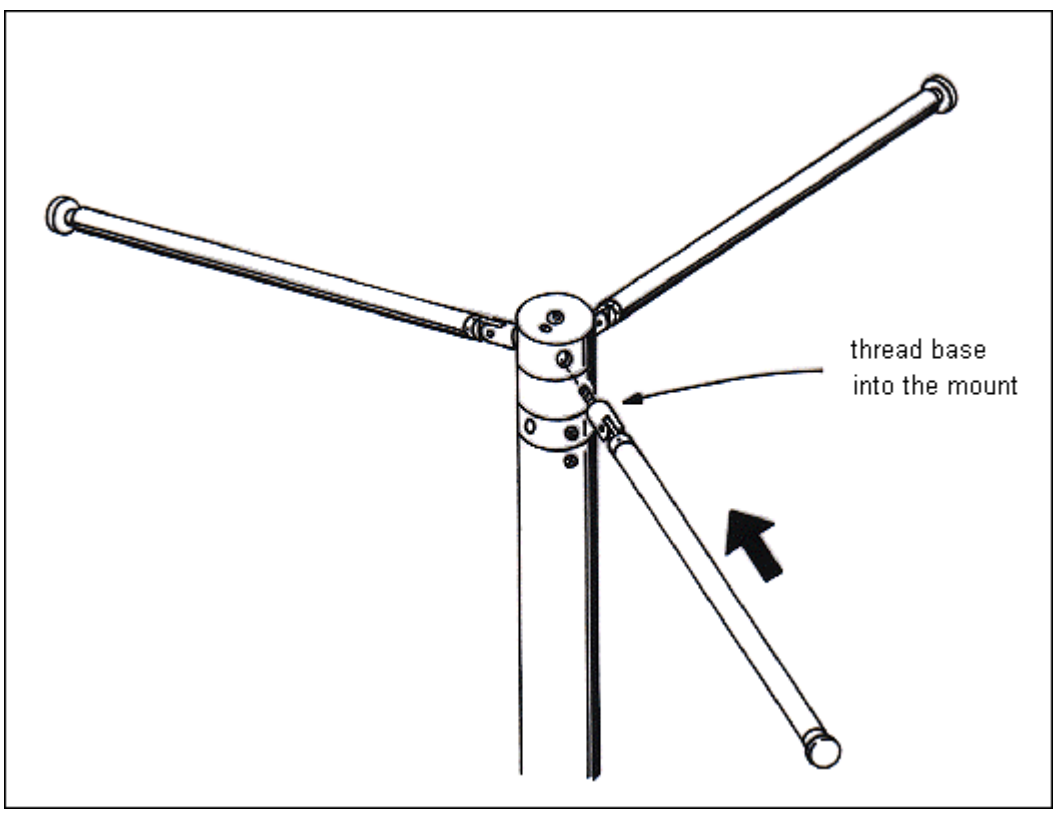


Figure 10.11 Telescopic antennas attached to the disc element mount.

Next, attach the three long telescopic antennas into the cone base mount under the first set of antennas (see Figure 10.12).

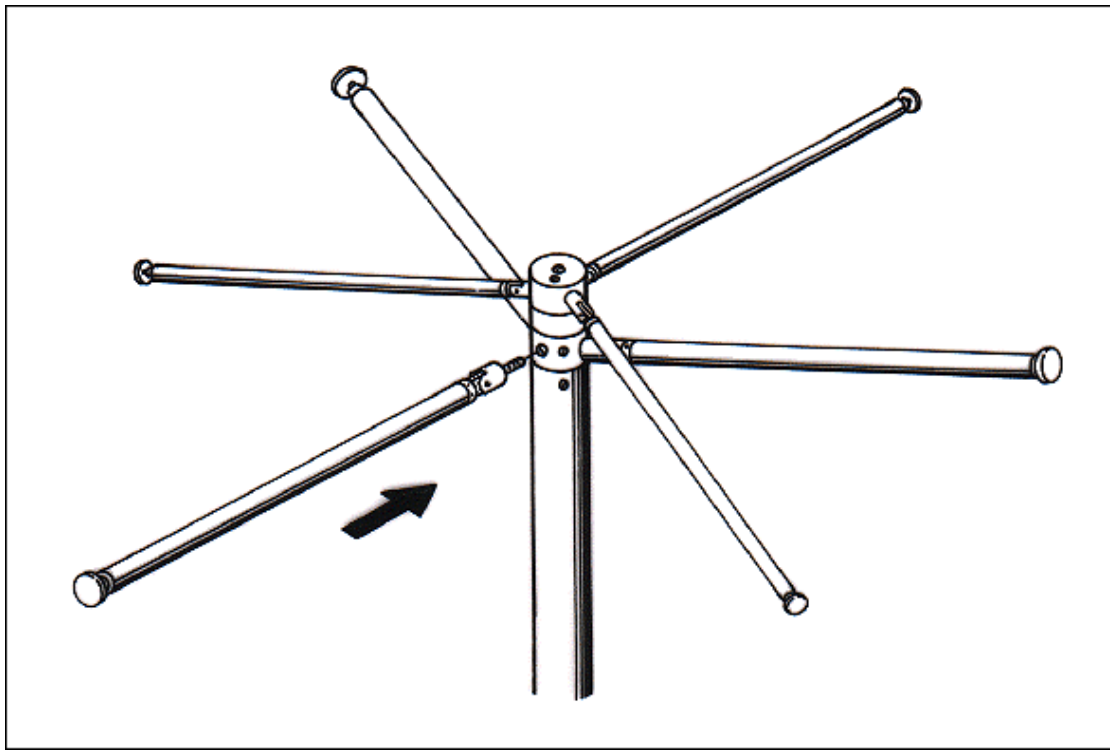


Figure 10.12 Attachment of long antennas to the cone element mount.

Attach the assembled antenna to the mast by using the aluminum mounting plate and one U-bolt. Extend the top telescopic antennas to their full lengths maintaining them in horizontal position. Similarly, extend the three long telescopic antennas to their full lengths but they must be bent to about 60 degrees angle drooping downwards to the ground. See Figure 10.13.

Figure 10.13 Mounting the completed antenna to the mast.

To carry the CD-2T in collapsed form for transportation, retract all the telescopic elements and bend them towards the mounting tube (see Figure 10.14).

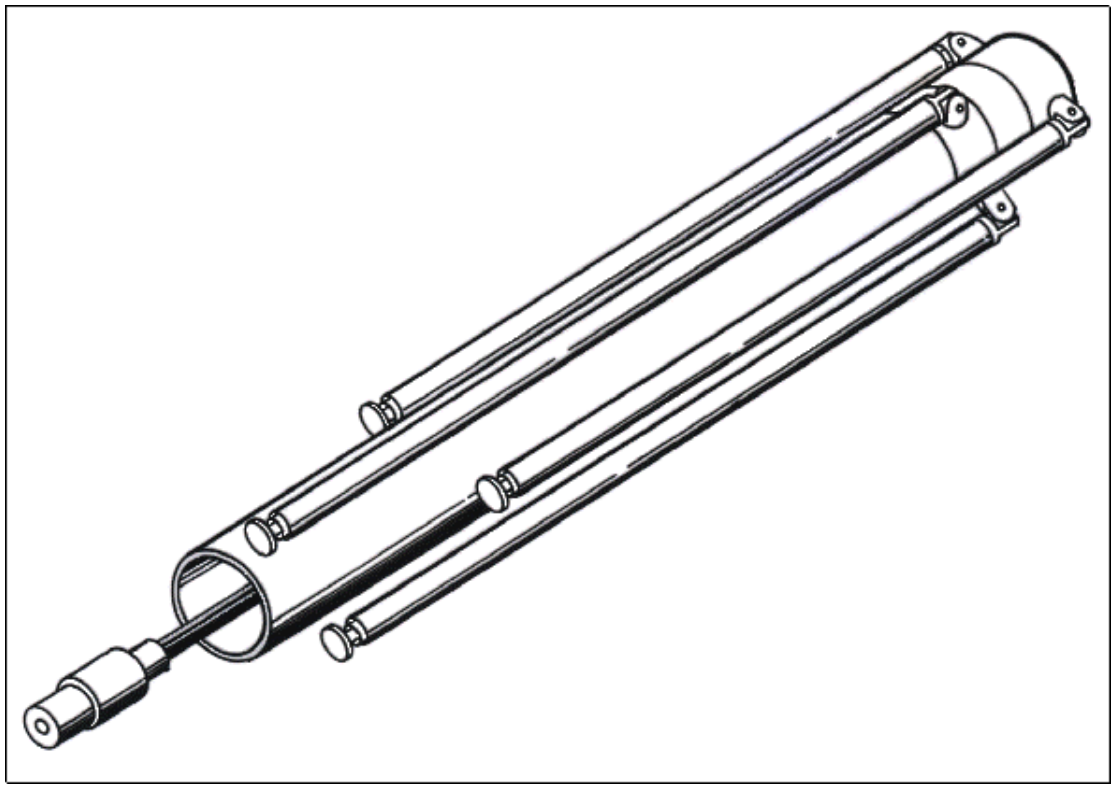


Figure 10.14 Antenna CD-2T in collapsed form.

11 5/8 WAVE ANTENNA

Model WA-2

Probably one of the most popular vertical antennas for both mobile and fixed station installations is the 5/8 wavelength vertical because it has some gain over a dipole. It is omni-directional and can be used either with radials or a solid-plane body (such as the one afforded by a car).

A version of a 5/8 vertical with radials is presented in this chapter. It is designed for fixed station installations. The common practice of radio operators is to install this antenna atop a tower with rotatable Yagi arrays positioned a few feet below it. The two antennas are connected to a common transceiver via a switching box. Only one antenna is active at one moment. The 5/8 wave vertical is used as a monitoring antenna because of its omni-directional characteristics. Once a contact has been established during operation, the operator quickly switches over to the Yagi antenna and beams it towards the other station to optimize communications. When the contact is finished the transceiver is again switched back to the 5/8 wave vertical antenna. This does not mean however that the average radio operator who cannot afford to erect a tower and a Yagi array should refrain from installing a 5/8 wave vertical. A properly constructed 5/8 wave vertical antenna if used singly works perfectly well!

Perhaps one advantage of constructing this antenna by the radio operator himself is the over-all cost of the unit. All of the materials used in this model are readily available at hardware stores and can be bought cheap. In comparison, a commercial version of this antenna costs more than a thousand pesos!

This antenna model WA-2 is designed to operate in the 140-150 MHz VHF band. It exhibits an SWR of less that 1.5:1 over the entire band if properly tuned. It has a gain of 1.8 dB over a standard dipole reference.

Scaled down appearance:

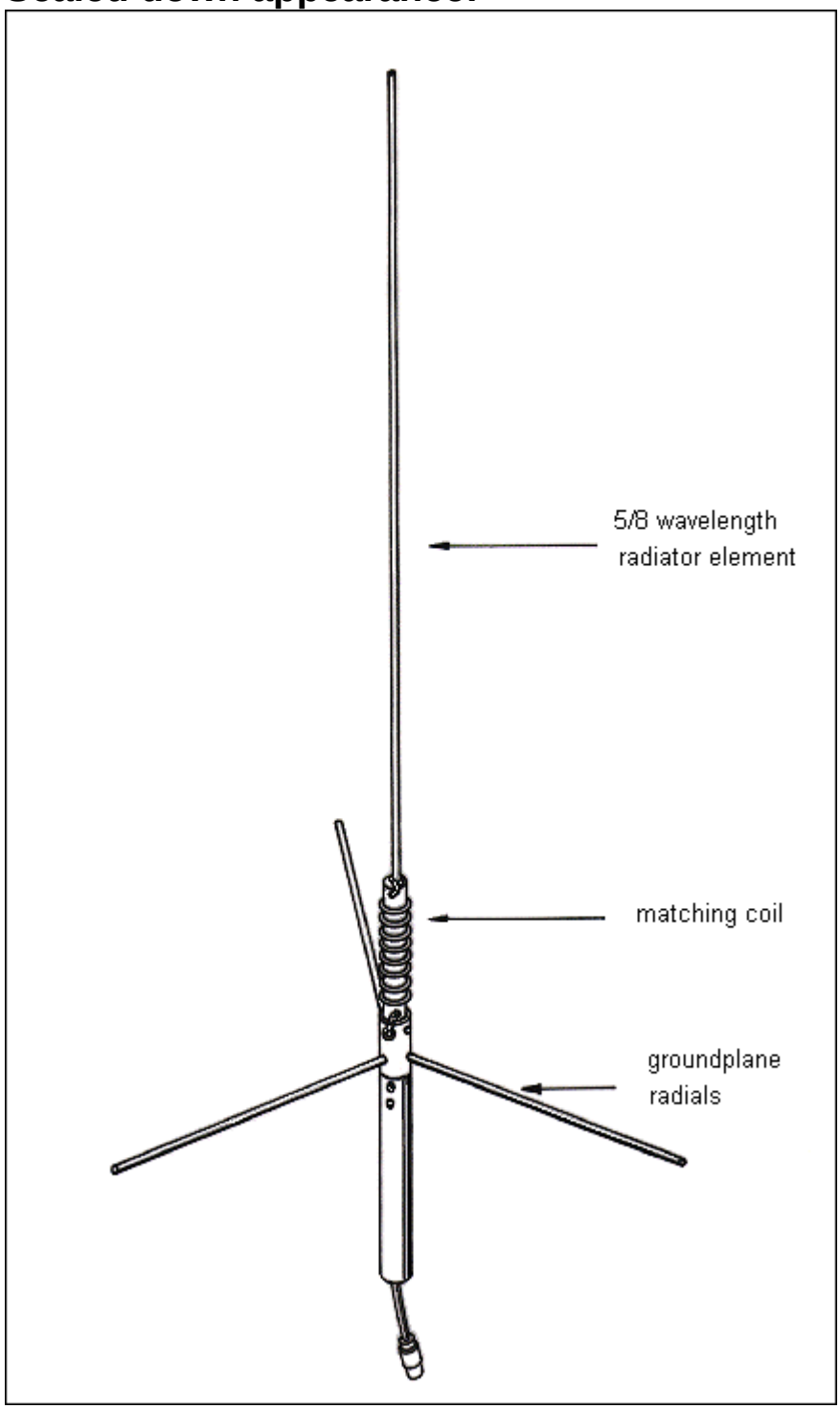


Figure 11.1

Materials List

Quantity	Specification/Description	Dimensions
2	Brass rods 1/8" diameter the brass rod for acetylene welding is recommended	
3	Brass rods 3/16" diameter	28 " long
1	Engineering plastic rod see text for dimensions	
1	PL-259 VHF connector	
1	PL-258 VHF straight connector	
1	Aluminum bushing - see text for dimensions	
1	Aluminum tube	1" id* x 8"
1	Copper wire gauge no. 14	20" long
1	Aluminum plate 1/8" thick	3" x 6"
4	U-bolts with accompanying hex nuts and lockwashers	
1	Coaxial cable RG-58/U	12" long
1	Stove bolt - brass or GI	1/8" x 3/8"
6	Self tapping metal screws	1/8" x 3/8"
1	Eye terminal vinyl insulated	1/16" id*
1	Short hook-up wire	3"- 4" long

* id - inside diameter

Construction

First prepare the plastic coil form by machining the engineering plastic rod to the dimensions shown in Figure 11.2.

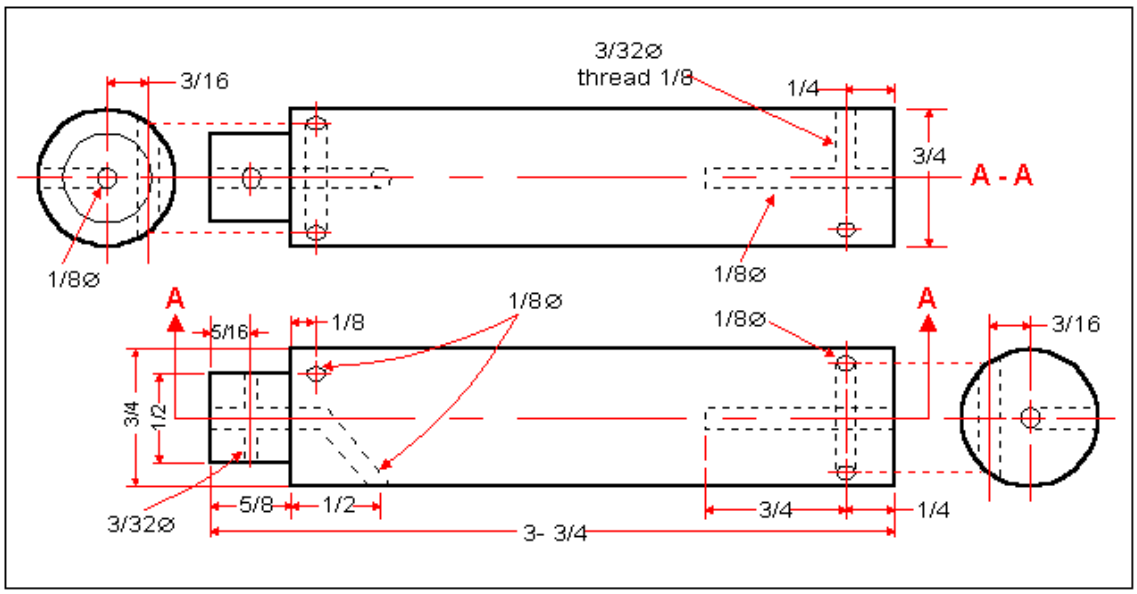


Figure 11.2 Coil form dimensions.

Next, prepare the bushing by machining the aluminum rod to the dimensions shown in Figure 11.3.

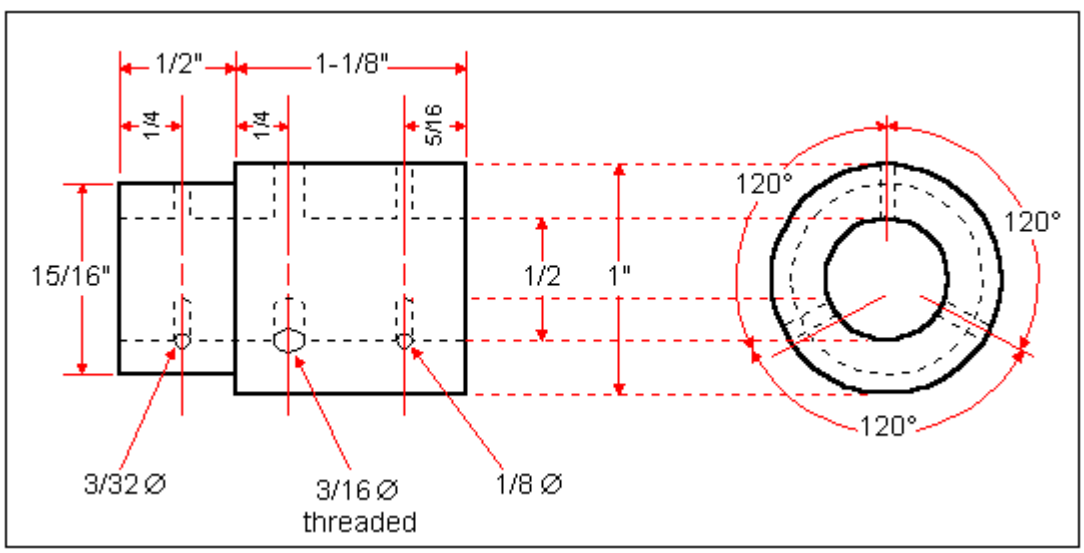


Figure 11.3 Aluminum bushing dimensions.

Assemble the plastic coil form and the aluminum bushing together. Secure the assembly with metal screws. Screw only through the two holes and temporarily leave the third hole unscrewed (see Figure 11.4).

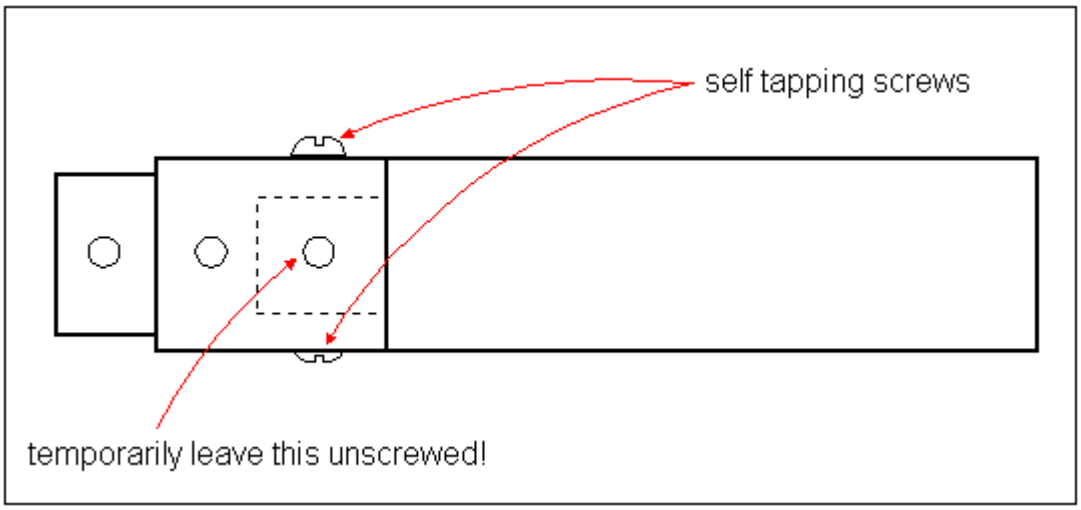


Figure 11.4 Assembling the coil form and the bushing together.

Prepare the radiator element. File a notch at one end of each 1/8" diameter brass rod. Join and solder the two notched ends together to make a long single rod (see Figure 11.5).

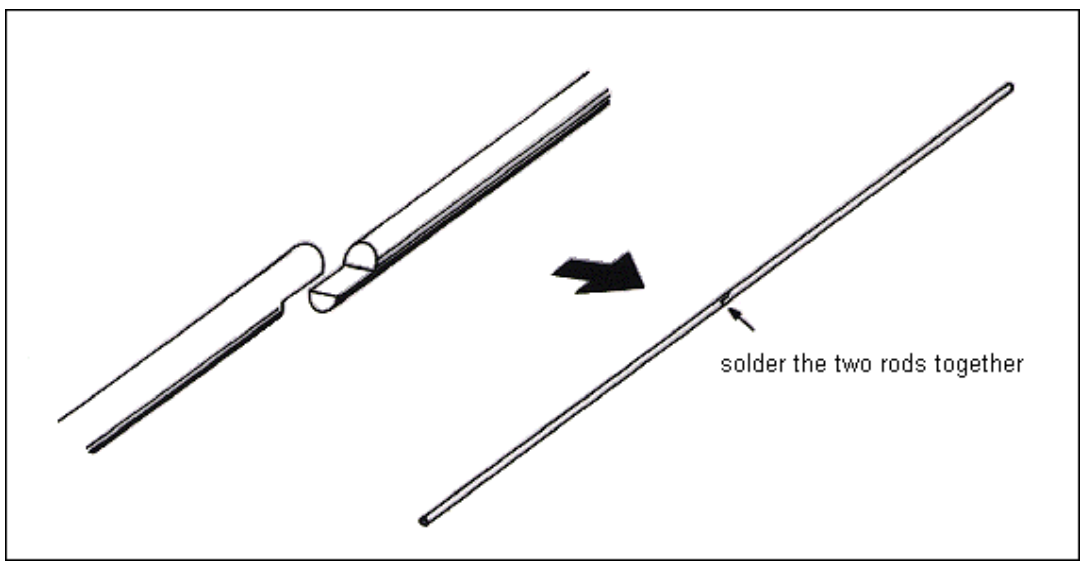


Figure 11.5 Joining the two brass together to make the radiator element.

Insert one end of the radiator rod into the top hole in the plastic coil form. Screw forcibly the 1/8" x 3/8" stove bolt through the hole at the side of the plastic coil form pressing the brass rod inside to hold it firmly (see Figure 11.6).

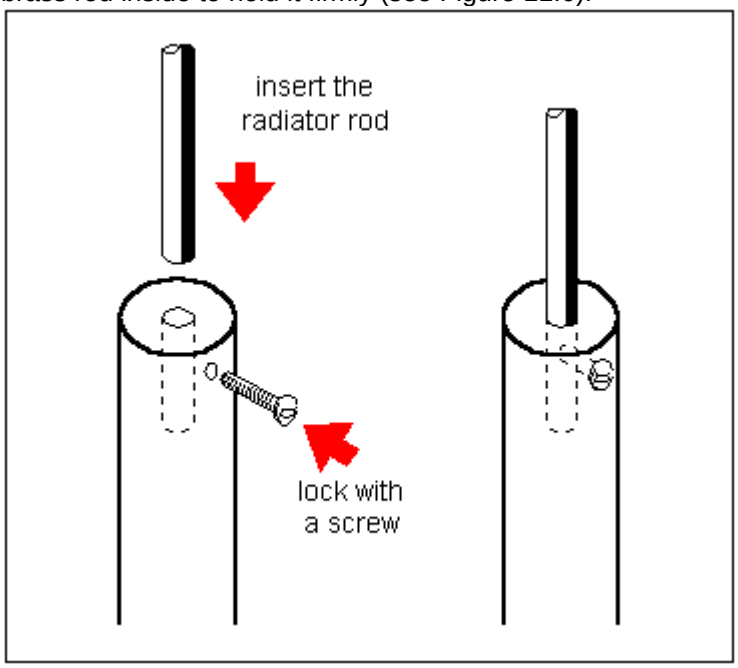


Figure 11.6 Securing the radiator element into the coil form.

Cut the brass rod to a length of 46 inches measuring from the point where it emerges from the plastic form (see Figure 11.7).

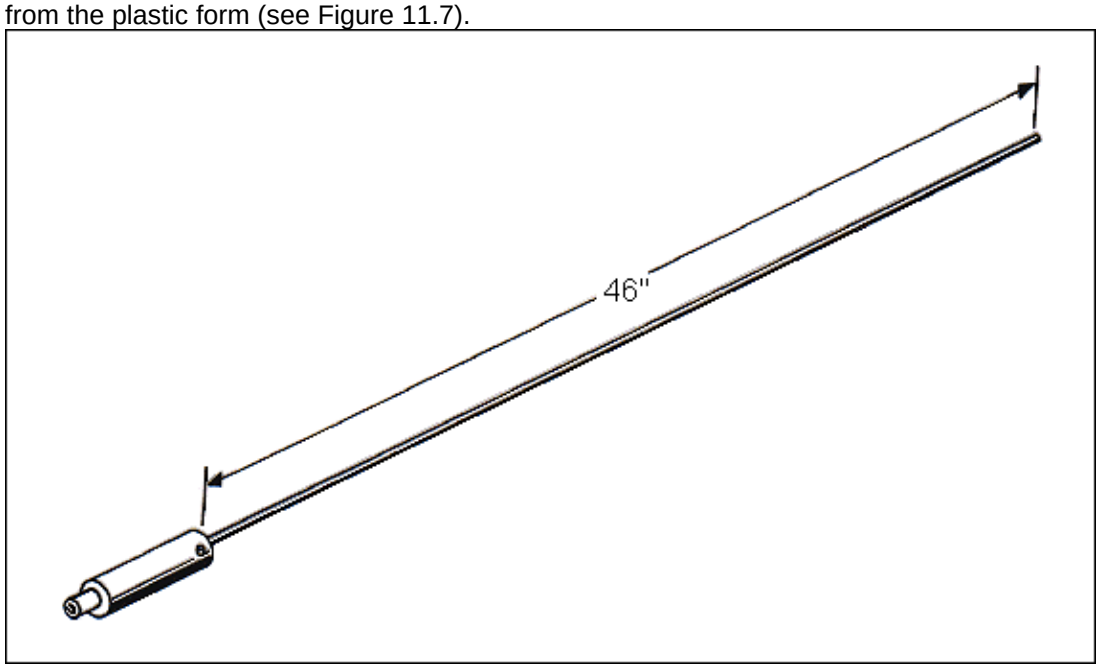


Figure 11.7 Cutting the radiator element to its proper length.

Wind the No. 14 copper wire around the coil form. Wind 10 and 1/2 turns evenly spaced and distributed to cover most of the length of the plastic form. Solder the top end of the copper wire to the base part of the brass rod (see Figure 11.8).

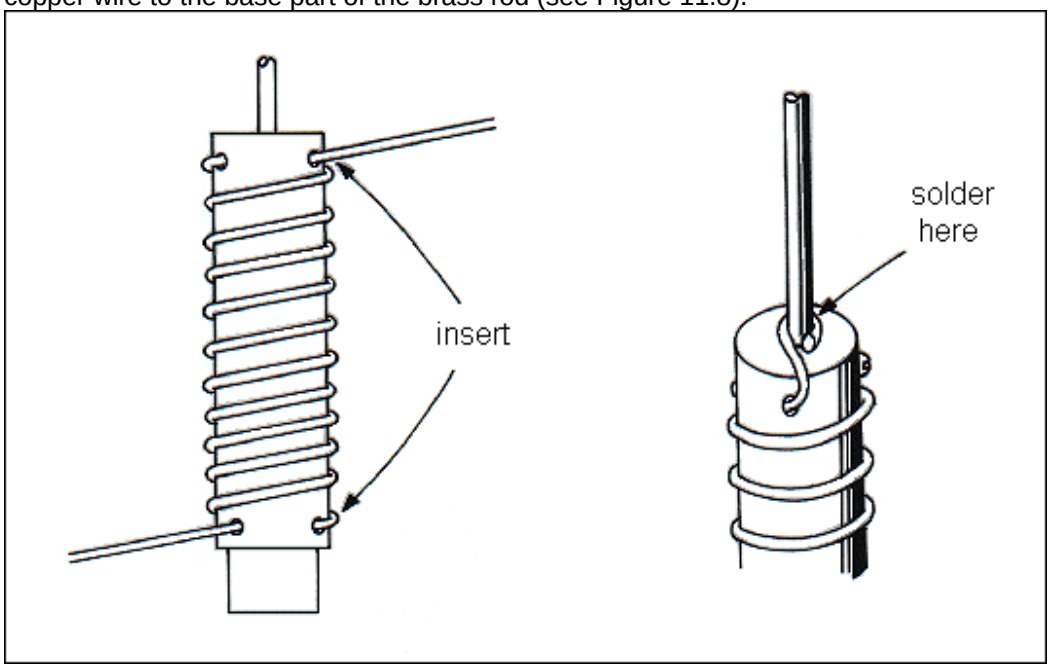


Figure 11.8 Winding the coil around the coil form.

Solder an eye terminal to the lower end of the copper coil. The eye terminal must be positioned in such a way that its eye is aligned with the unscrewed hole in the aluminum bushing. After you have soldered the eye terminal, attach it into the aluminum bushing with a metal screw (see Figure 11.9).

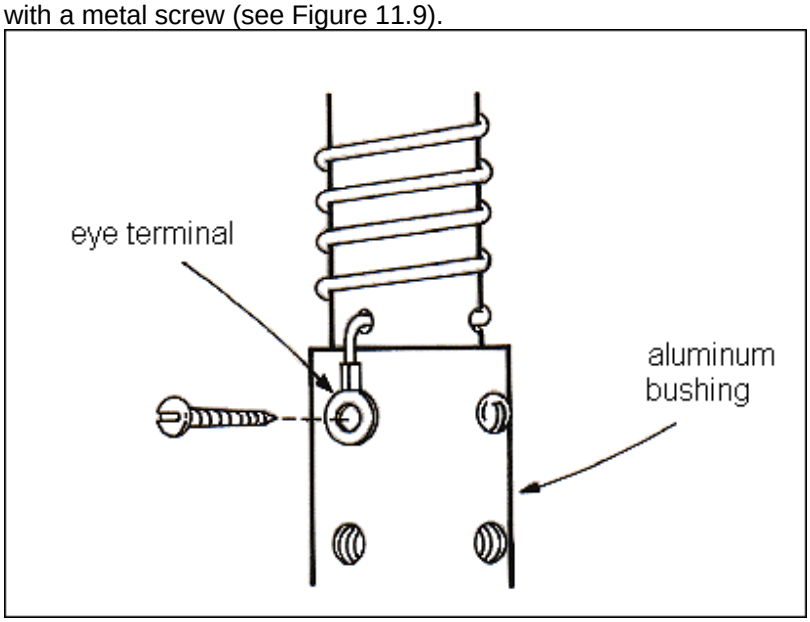


Figure 11.9 Securing the coil to the aluminum bushing.

Cut a short length of stranded hook-up wire (about 3 inches). Insert it into the hole in the plastic coil form until it protrudes from the center hole at the bottom. Solder the upper end of the hook-up wire to approximately 6 and 1/2 turns counting from the coil's lower end connected to the aluminum bushing (see Figure 11.10). This connection is temporary only and it may be necessary to move the wire during tune-up procedure.

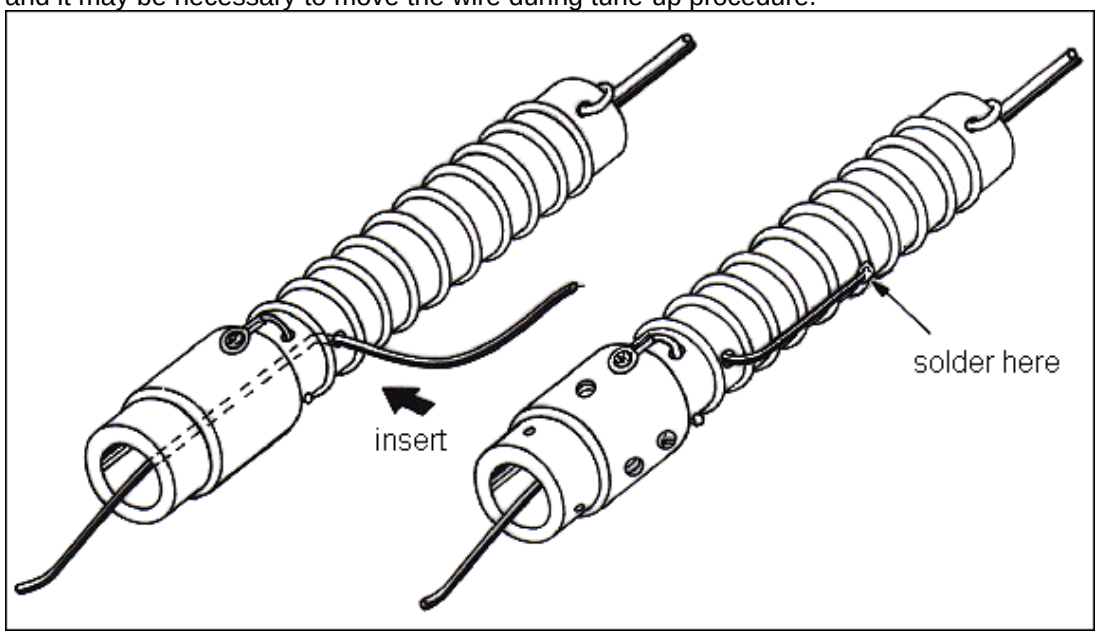


Figure 11.10 Tapping the coil for feedpoint.

Cut 12" length of coaxial cable RG-58/U and separate the braid from the inner conductor at one end (make a pig tail). Solder the inner conductor into the hanging end of the hook-up wire at the bottom of the plastic form. After joining the two wires, insulate the joint either with a shrinking tube or just a plain vinyl tape. Solder an eye terminal into the braid of the coax cable (see Figure 11.11).

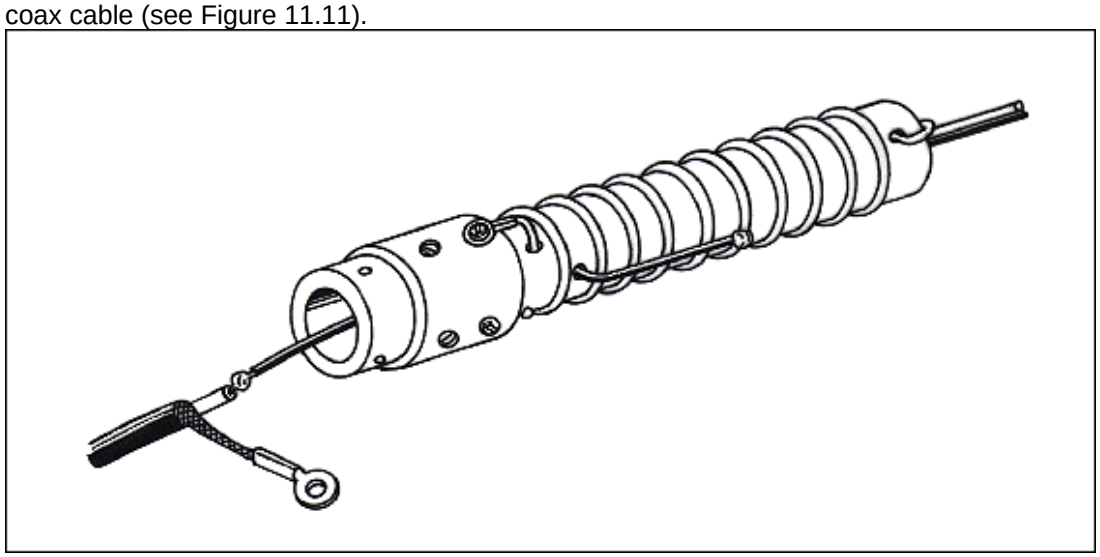


Figure 11.11 Connecting the coaxial cable to the hook-up wire.

Next step is to prepare the mounting tube. Cut 1" diameter tube to a length of 12 inches and drill three holes at one end. The holes must be 1/8" in diameter and equally spaced from each other. Drill a single hole at the same end but slightly below one of the first 3 holes (see Figure 11.12).

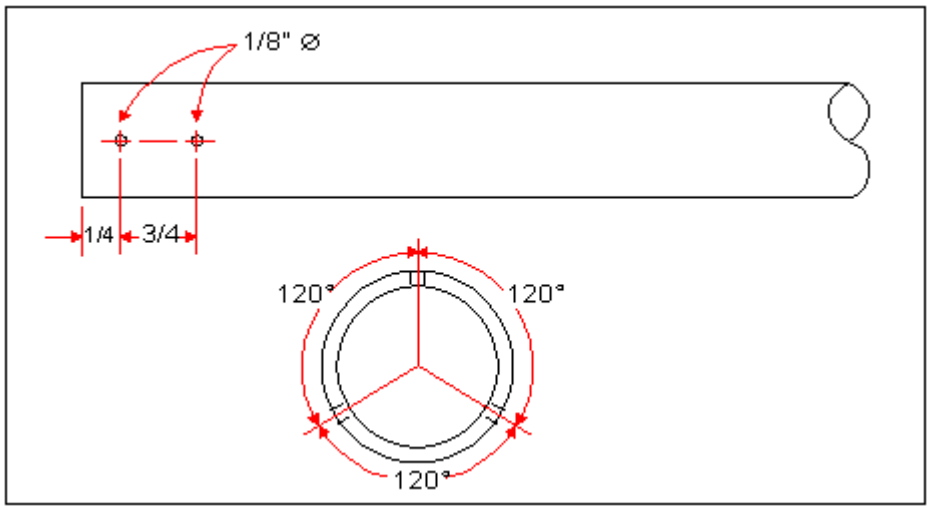


Figure 11.12 Preparing the mounting tube.

Next, insert the free end of the coaxial cable into the mounting tube starting at the end with sideholes. When the aluminum bushing and tube meet, insert the bushing inside the tube and align the holes at their sides. Secure the bushing into the tube by screwing self tapping screws into the holes (see Figure 11.13).

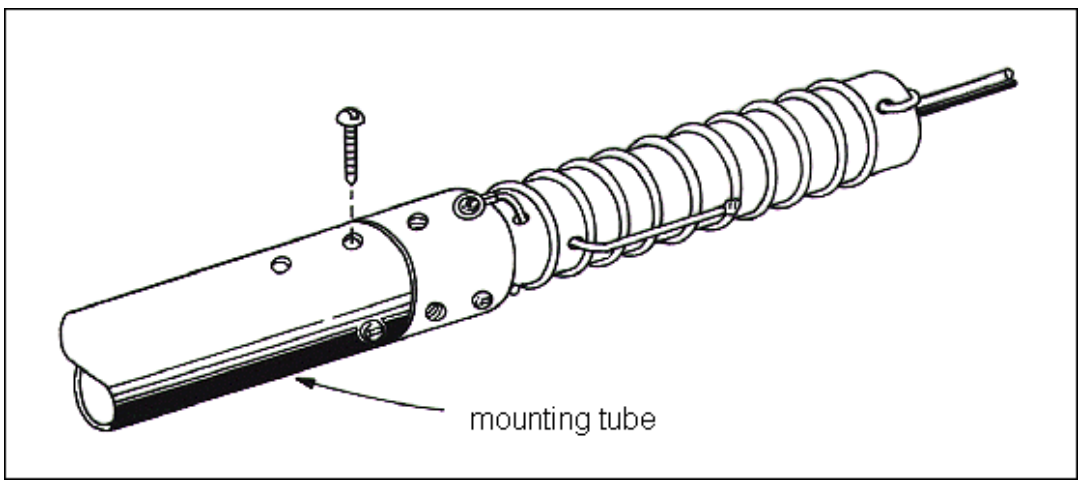


Figure 11.13 Securing the aluminum bushing into the mounting tube.

Secure the braid by inserting a self tapping screw into the lone hole and tapping into its eye terminal inside the tube. Tighten the screw to hold the eye terminal against the wall of the tube (see Figure 11.14). You may need to use a wooden stick inserted into the tube to position the eye terminal exactly under the hole.

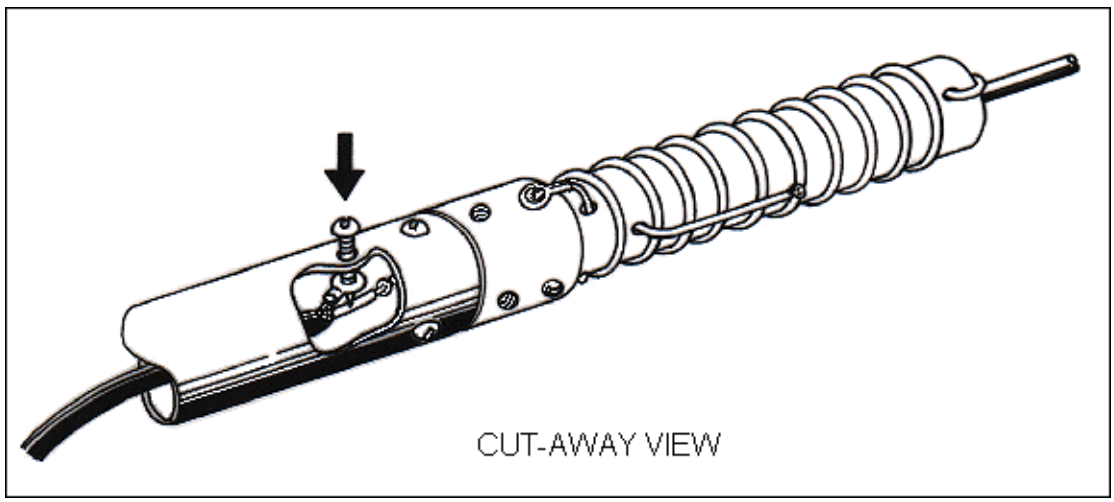


Figure 11.14 Securing the braid inside the tube with a metal screw.

Next, prepare the groundplane radials. Cut three lengths of 3/16" diameter brass rods. Note that these rods are larger than the radiator rod. Each rod must be 28 inches long and threaded at one end (see Figure 11.15).

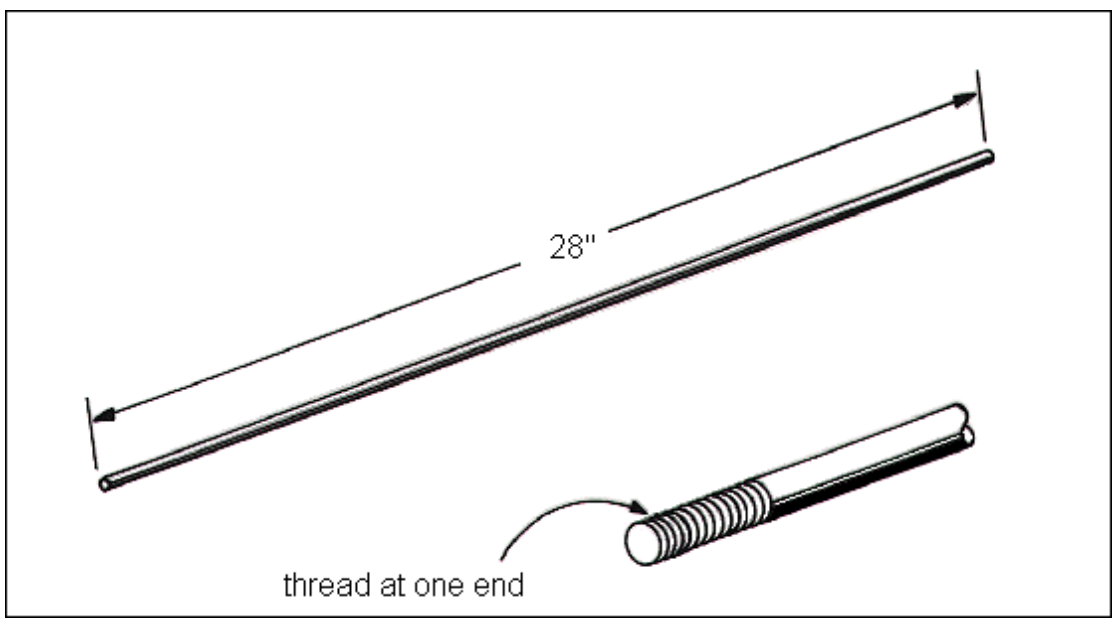


Figure 11.15 Cutting the radials and threading one of their ends.

Final assembly and installation

Attach the three groundplane radials into their mounting holes at the aluminum bushing. Mount the antenna to the mast by using a metal plate adaptor similar to one described in the preceeding chapters (see following illustration).

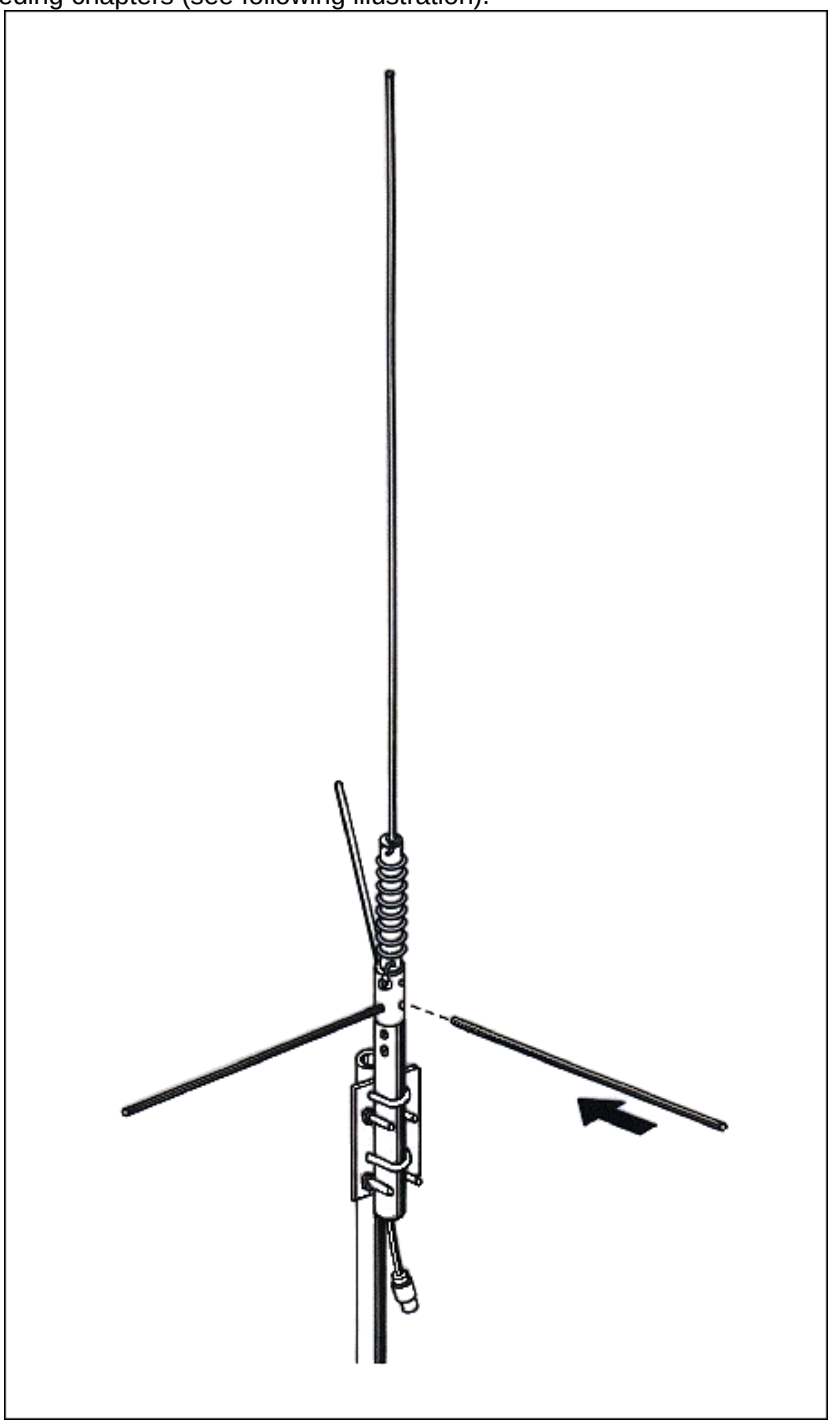


Figure 11.16 Mounting the WA-2 to the mast.

Tuning WA-2 to resonance

Mount the antenna to the mast as previously described. Connect a coaxial cable into the PL-258 connector of the antenna and attach the other end into the output of an SWR meter (marked with 'antenna'). Attach also a short coaxial feeder into the input of the SWR meter (usually marked 'transmitter') and the other end of the feeder must be plugged into the output connector of your transceiver (see Figure 11.17).

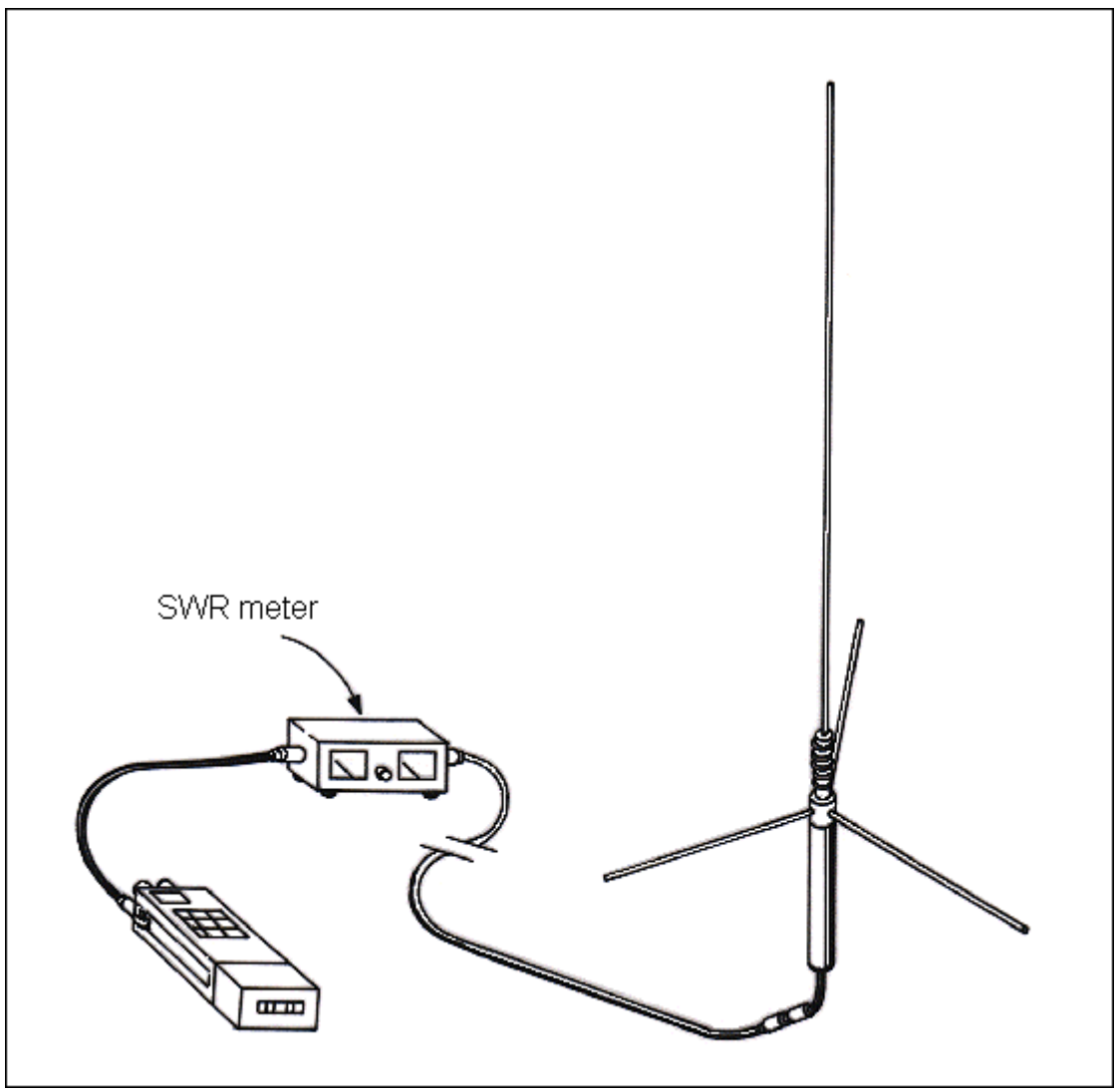


Figure 11.17 Preparing the antenna for tuning to resonance.

Set your transceiver's frequency to the center of the band and key the PTT. Read the SWR response and write it down to a chart similar to the one shown in Figure 11.18.

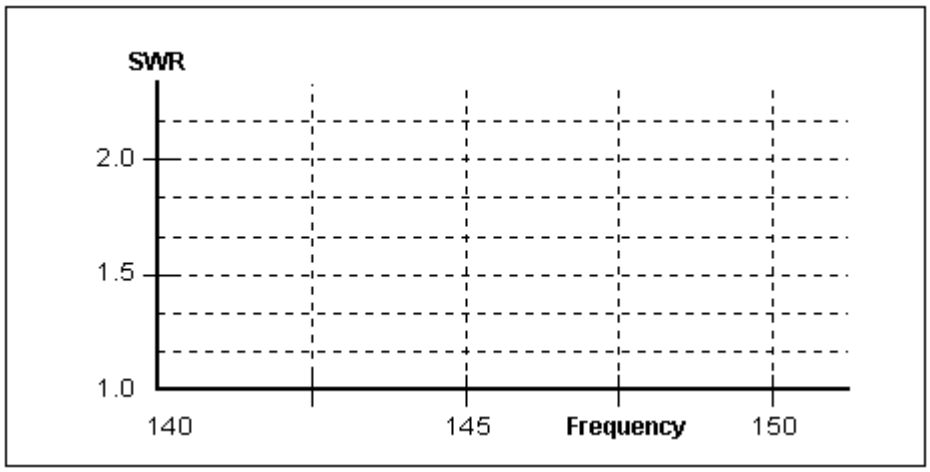


Figure 11.18 SWR chart.

Resolder the hook-up wire to another point in the copper coil to get the lowest SWR response in the center frequency and a relatively flat response over the entire band similar to the charted response shown below (see Figure 11.19).

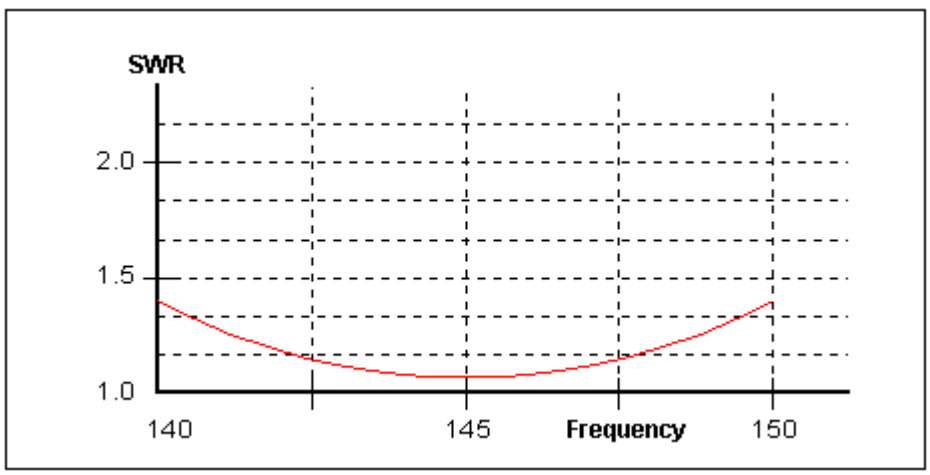


Figure 11.19 A sample of a charted SWR response.

You can move the soldered point or tap either way - left or right - depending on how the SWR responses. If you have moved the tap to the right and the SWR went higher then obviously you must move the tap to opposite direction - to the left. You must check the SWR reading over the entire band everytime you move the tap. Move the tap only about 1/4 inch farther each time. After you have found the best point in the copper coil, solder the hook-up wire permanently (see Figure 11.20).

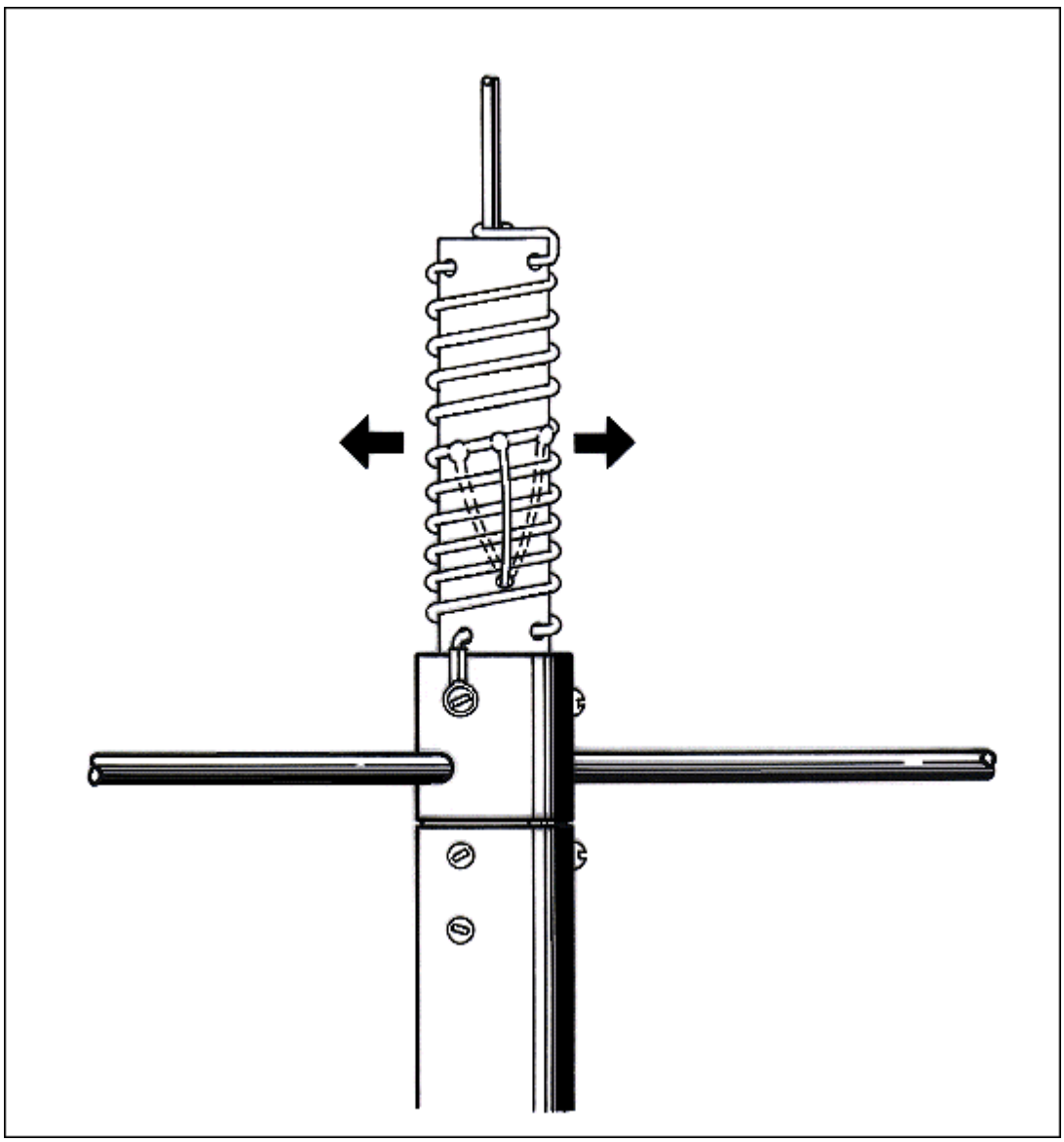


Figure 11.20 Resoldering the tap to a different point to find the best SWR response.

Dismount the antenna from the mast and remove its three groundplane radials. Place the heat shrinking tube into the antenna wrapping the entire coil form and heat it over a flame or with a blow drier. The coil form and shrinking tube must be rotated continously over the heat to result to an even shrinking of the tube (see Figure 11.20). If you are heating the tube over the flame, don't let the flame touch the tube directly.

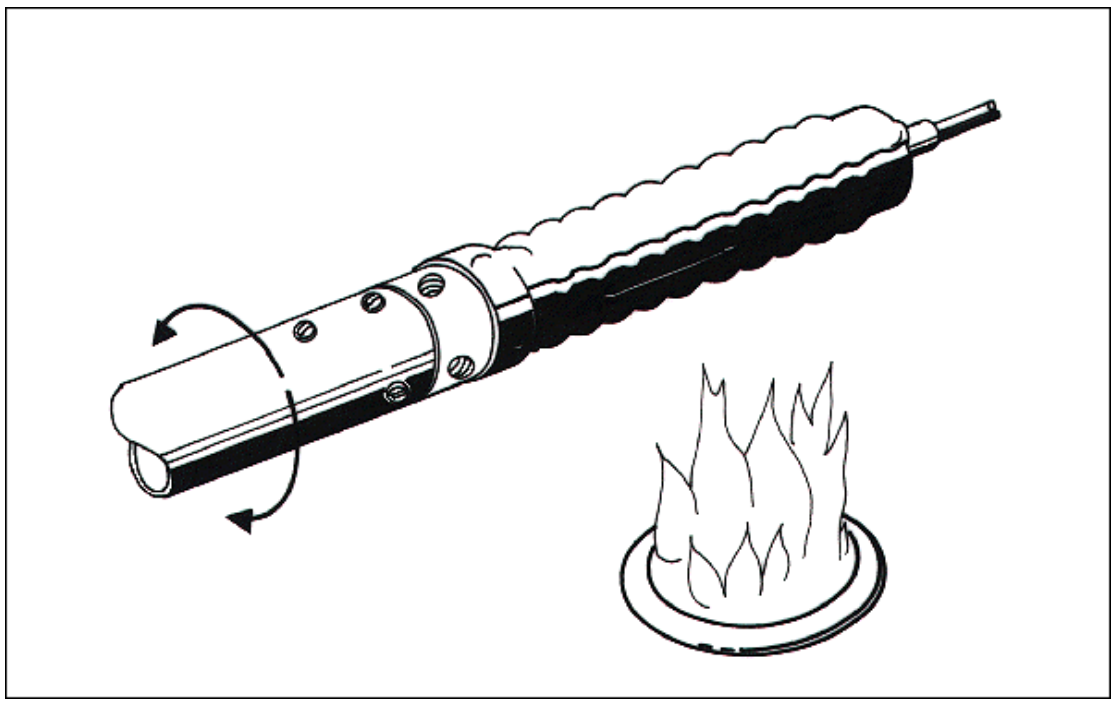
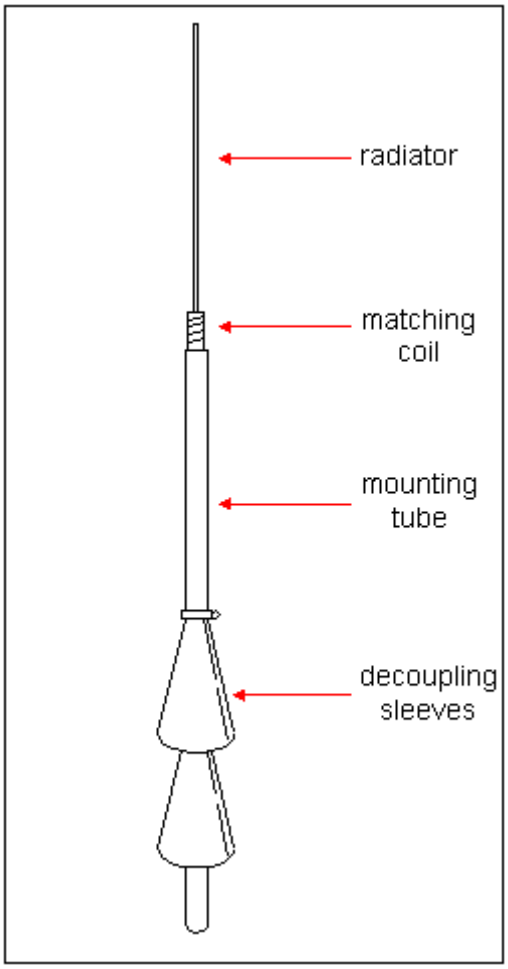


Figure 11.21 Heating the shrinkable tube.

12 5/8 WAVE ANTENNA

Model WD-2



This antenna is an improvement of the basic design of 5/8 wave vertical with radials. As can be clearly seen in the following illustration, it has two metallic cones attached to a long tube which doubles as support for the radiator element. The cones are not intended for novelty but serves a very important purpose for a more efficient performance of the entire antenna system. Its function is to nullify the unbalanced coupling between the transmission and the antenna feedpoint and prevent the unwanted current from flowing on the outside of the coaxial cable.

Why is this so?. Well, let us first go back to some basics to understand this phenomenon. In a perfectly balanced antenna the electrical current within each leg of the element is symmetrical. There will be no problem in coupling the RF signal to its feedpoint when a balanced feedline is used. However, if a coaxial cable is used to feed the antenna, the coupling action is inherently unbalanced because of the physical construction of the coaxial cable. Stated simply, the outside part of the outer conductor is not coupled to the antenna in the same way as its inner part is coupled to the inner conductor. The overall result is that current will flow on the outside of the outer conductor. This current is negligible in the HF frequencies but must not be ignored in VHF or UHF frequencies. This problem is remedied by the metal cones described in this particular model - it detunes the system for stray currents present on the ouside of the line. The cones are also called " detuning sleeves" or "decoupling sleeves".

An antenna system with a properly decoupled line is commonly used in repeater systems because by the very nature of its design, a repeater station is very sensitive to any kind of stray RF signal. A repeater station has both receiver and transmitter units simultaneously operating when used. Although the frequency of the transmitter unit is different from the frequency of the receiver unit, the very close proximity of the two units tends to blank out the generally weak signals from distant stations. This results to a phenomenon called "desensitation" or "desense" where the repeater cannot receive the signals from the user stations.

Feedback also results to loud squeal heard by the users. The entire system ceases to function as a repeater then. Desense and feedback is avoided by using high-Q cavitly filters inserted in the transmission line for the transmitter or receiver antenna or both.

Additionally, the automatic switching electronics of the repeater is also protected against picking up unwanted RF by enclosing it in a metal box and by extensive use of decoupling circuits in all the leads going in and out of the box. However all of these efforts could fail if the stray current that travels along the outside part of the transmission line is so strong that it penetrates all filters installed in the repeater system. Using a decoupled antenna system such as the one described in this chapter will save you from the trouble.

The model WD-2 is specifically dimensioned to operate in the 140-150 MHz band. It exhibits an SWR response of less than 1.5:1 over the entire band. The radiation pattern is omni-directional. It has a gain of 1.8 dB compared to a standard dipole reference.

Materials needed

This antenna is basically the same with the antenna model WA-2. The difference between the two models is that the mounting tube for the model WD-2 is 104 inches long and has no groundplane radials but instead it has two decoupling sleeves made of metal cones attached to the lower portion of the mounting tube.

Construction

Construct the antenna following the procedures described for the model WA-2 in chapter 11 except for the length of the mounting tube which is 104 inches long for the model WD-2. Also skip the procedure for preparing the groundplane radials, you don't need them for this antenna anyway. Furthermore, before you tune the antenna to resonance, construct the decoupling sleeves and attach them to the mounting tube following the procedures described here.

Cut the cone form from a metal plate (GI sheet or aluminum) following the dimensions shown in Figure 12.2. In forming the cone, overlap its edges and drill holes along the edge. Rivet the ovelapping edges through these holes.

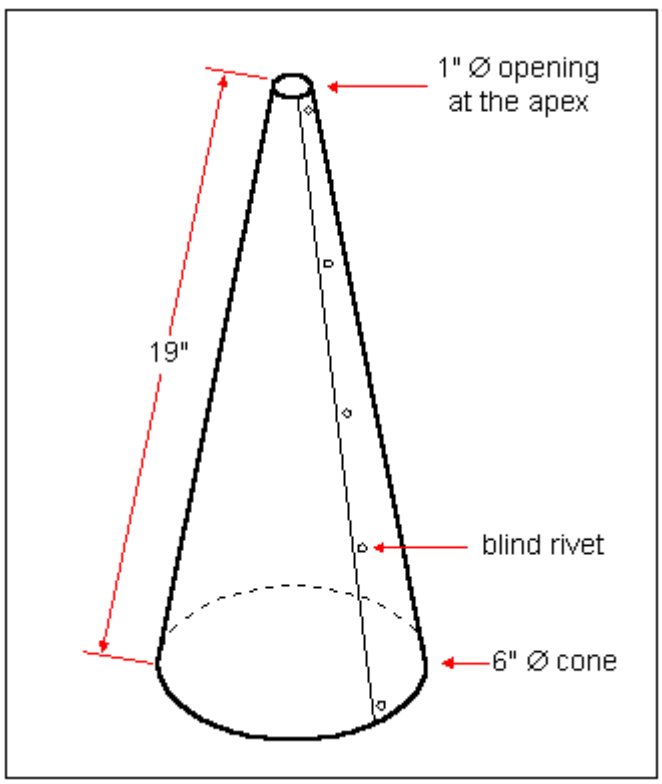


Figure 12.2 Fabricating the decoupling sleeve (or 'decoupling skirt').

Cut a narrow strip out of a similar material and form it to a ring with a diameter of 1 inch as shown. Leave a small gap between the two ends. This ring will serve as a mounting sleeve so that the decoupling sleeve or skirt can be securely clamped to the mounting tube (see Figure 12.3).

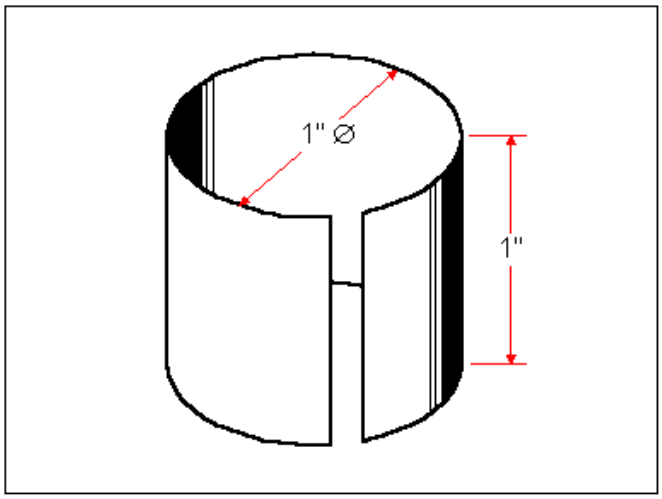


Figure 12.3 Preparing the metal ring.

Solder the ring to the apex of the cone (see Figure 12.4). If you use an aluminum plate you must electrically weld the two pieces together using a special welding technique with protective gas.

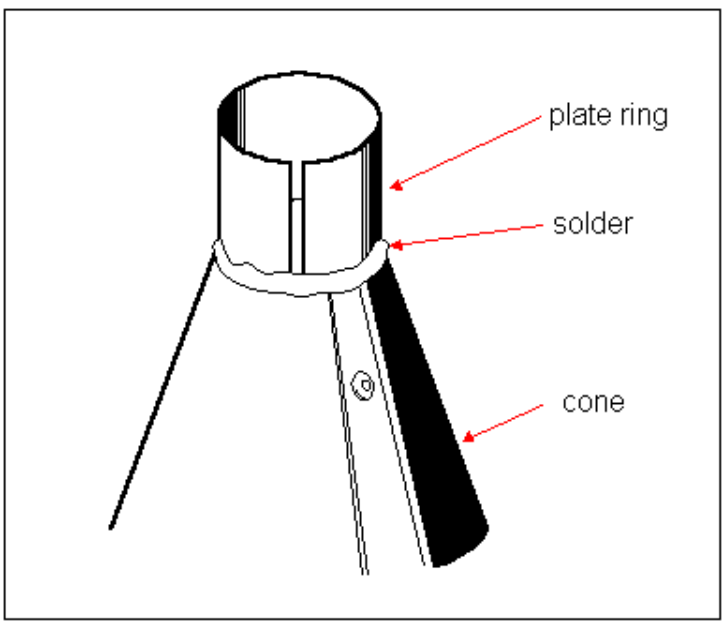


Figure 12.4 Soldering the ring to the cone.

Attach the two decoupling sleeves/skirts into the mounting tube following the distances shown in Figure 12.5. Place a tube clamp over each cone and tighten it to secure the cones firmly to the mounting tube (see Figure 12.6).

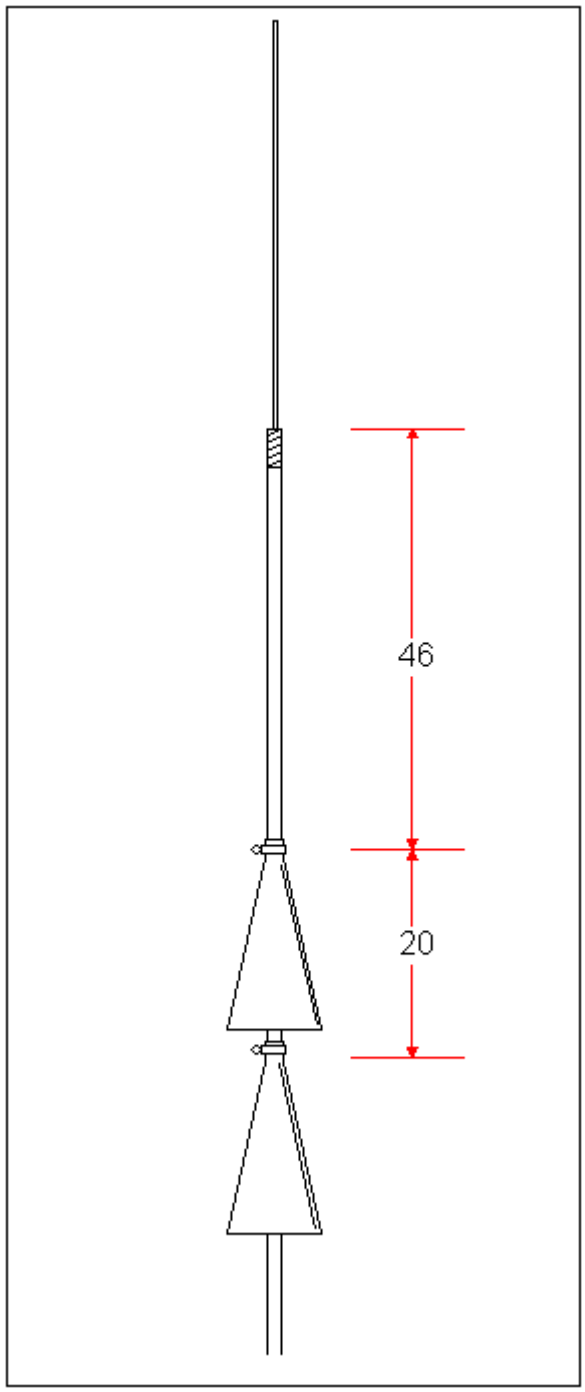


Figure 12.5 Mounting the decoupling sleeves to the antenna.

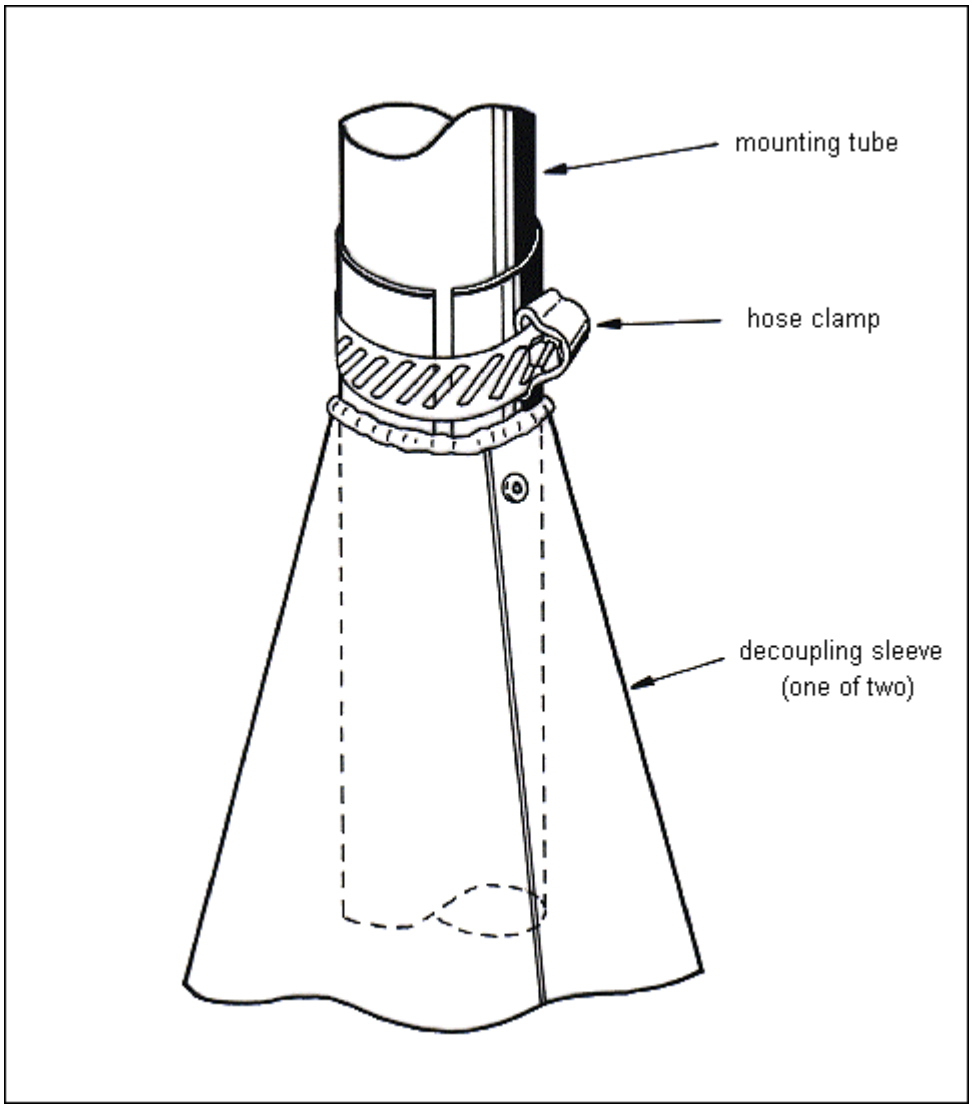


Figure 12.6 Securing the cone to the mounting tube with a hose clamp.

Tuning the antenna to resonance

The tuning procedure for this antenna is the same with the procedure for tuning the antenna model WA-2. Just follow the procedures described in chapter 11.

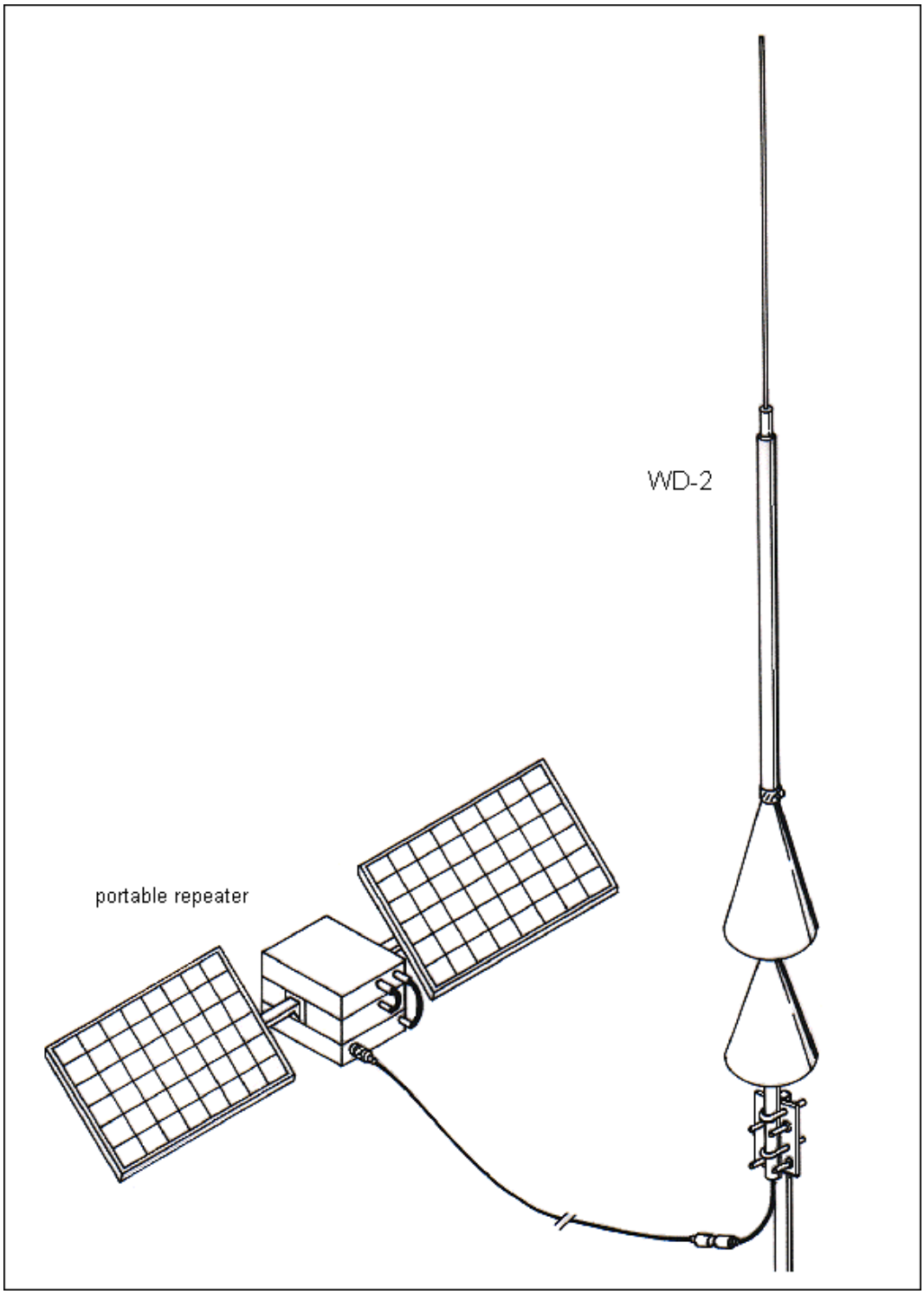


Figure 12.7 Mounting the WD-2 to the mast.

13 5/8 WAVE ANTENNA

Model PF-2C

Most mobile operators use portable handheld transceivers because these are lightweight and small. There are also available models today that equal the capabilities of their base station versions in terms of frequency coverage, sensitivity, computerized funtions, PLL stability and many other unique features. However, portable transceivers in general have low power transmitters because of obvious limitations in the type of batteries practically allowable for mobile operations. The average transmitting power of handheld units range from 0.5 watts to 5 watts maximum. Because of this, most antennas used for portable sets are of gain type to increase the effective radiated power.

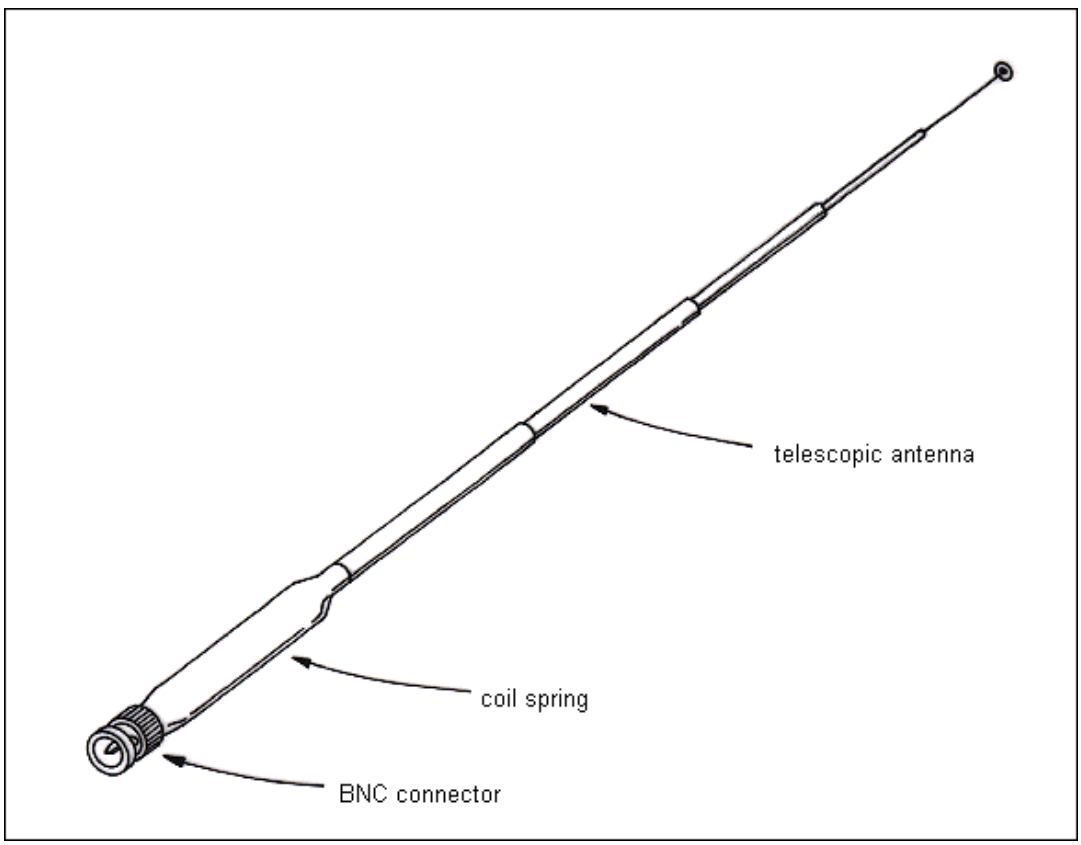


Figure 13.1 5/8 wave antenna model PF-2C.

The antenna described here is a portable version of a 5/8 wave vertical antenna. As stated earlier, an antenna of this length has a slight gain over a dipole. Approximately, a gain of 1.8 dB can be attained with this type of antenna. The radiator element of this model is made of telescopic rod so that the overall length of the antenna can be reduced if desired. It is loaded at the base by a coil that doubles as a flexible spring supporting the telescopic rod. The telescopic element may be used while retracted or collapsed and will function like an ordinary "rubber ducky" antenna that comes as a standard accessory for portable transceivers. Gain can only be realized if the antenna is used while the radiator is extended to full length.

Sometimes it is desirable to raise the height of the antenna to increase its effective range. Installing the antenna to a higher position clears it from most obstructions such as houses or trees and extends the horizon farther away thereby increasing the area covered or "seen" by the antenna. This can be accomplished by using a length of coaxial cable to connect the antenna to the transceiver. The antenna is then mounted high up in a post or tower. It can also be hung under a tree by using a non-metallic material such as nylon or fishline.

This particular model is dimensioned to operate in the frequency band of 140-150 MHz. It exhibits an SWR of less than 1.5:1 over the entire band if properly tuned. Tuning is easy as described in this chapter. The materials used for this model can be bought cheap and constructing it can save a lot of money. The total cost of the antenna is a mere fraction of the price of its commercial version. Furthermore an invaluable knowledge can be gained during the actual construction of this antenna.

Materials List

Quantity	Specification/Description	Dimensions
1	Telescopic antenna - approx. 7/16" od* of the base tube	46" fully extended 6" to 7" retracted
1	Brass wire no. 26 or approx. 3/32" diameter	24" or 60 cm.
1	BNC VHF male connector	
1	10 pF/150 Volts capacitor glass type	
1	Hook-up wire no. 22 stranded	4" long
1	Heat shrinkable tube	5/8" or 3/4" od* x 4" long

Miscellaneous: Epoxy glue

- od - outside diameter

Construction

Wind the brass wire into a spring like coil form. Wind 13 turns of the wire with a pitch of approximately 4 turns per inch. The total length of the finished coil is approximately 3 inches (see Figure 13.2). The inside diameter of the coil spring must be force fit to the outside diameter of the BNC connector or approximately 3/8" id*.

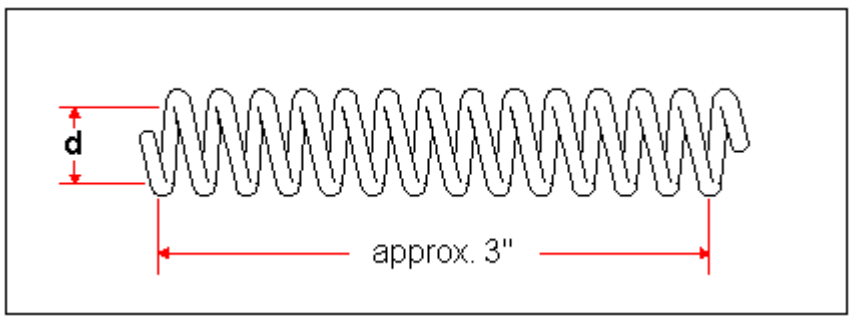


Figure 13.2 Constructing the spring coil.
*id-inside diameter

Solder about 2" long hook-up wire to the center pin of the BNC connector.

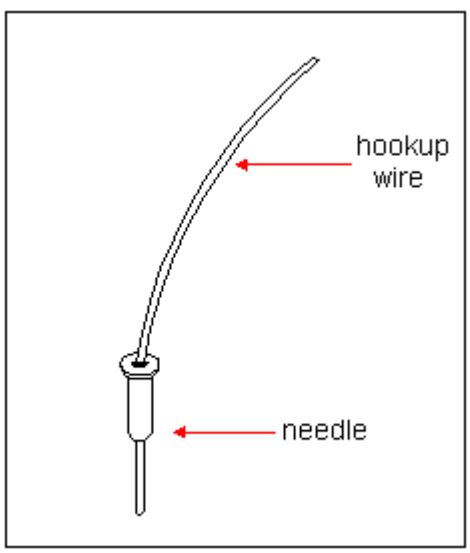


Figure 13.3 Soldering the hook-up wire to the center pin of BNC.

Place a moderate amount of epoxy glue around the soldered part of the needle. Avoid coating the epoxy around the body of the center pin. Insert the needle into the BNC connector and cover the empty space inside with a liberal amount of epoxy (see Figure 13.4). Let the epoxy cure and harden.

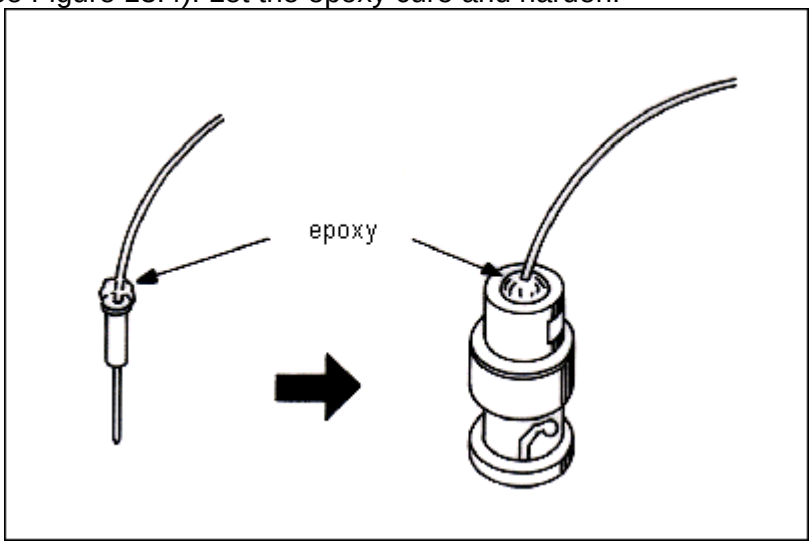


Figure 13.4 Fixing the center pin to the BNC connector with epoxy glue.

When the epoxy hardens insert the BNC connector into one end of the spring coil. Solder the part of the coil that wraps around the body of the BNC connector (see Figure 13.5).

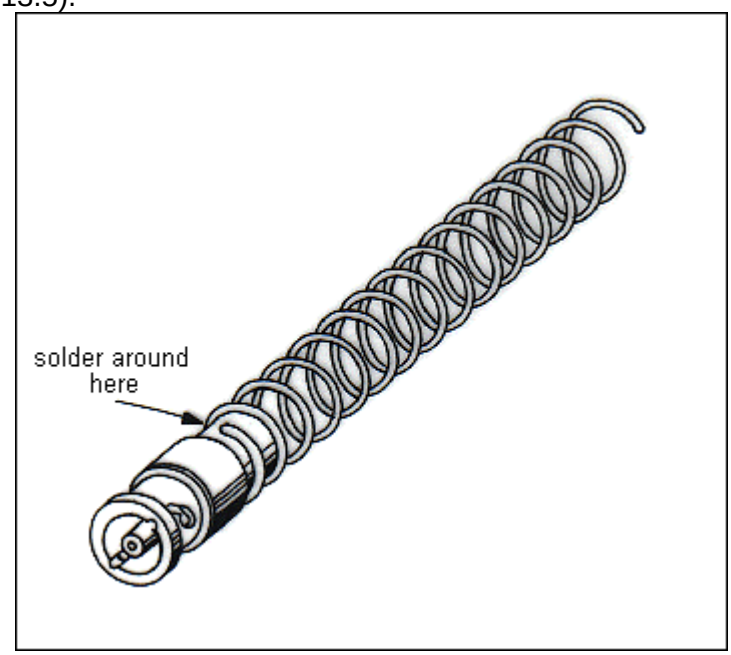


Figure 13.5 Assembling the BNC connector and the spring coil together.

Pry out the free end of the hook-up wire inside the coil spring and solder it to the point of the coil which is 1 and 1/2 turns counting from the ungrounded portion of the coil (see Figure 13.6).

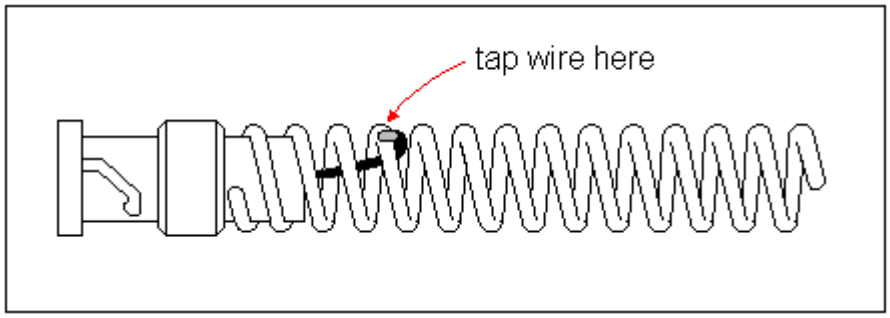


Figure 13.6 Soldering the hook-up wire to a temporary tap point.

Insert the 10 pF capacitor inside the spring coil and solder its lower lead to the grounded portion of the coil (see Figure 13.7).

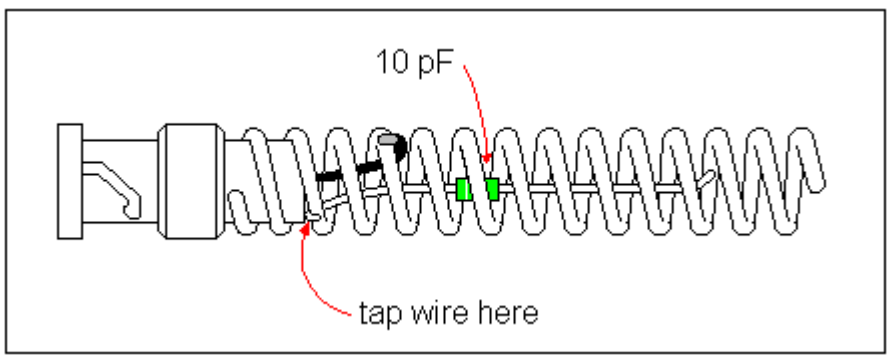


Figure 13.7 Soldering one lead of the capacitor to the grounded portion.

Next, solder the upper lead of the capacitor to the 6th turn of the coil spring counting from the ungrounded portion. See Figure 13.8.

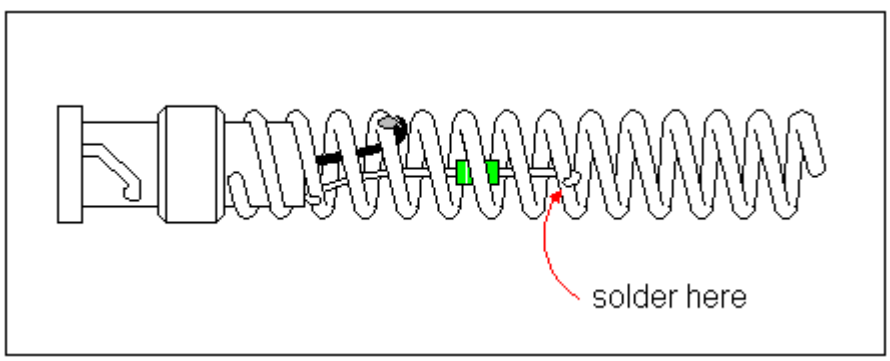


Figure 13.8 Soldering the upper lead of the capacitor.

Next step is to insert the base of the telescopic antenna into the open end of the coil spring. Let 2 turns of the coil hold the base of the antenna and solder it to secure the two pieces together (see Figure 13.9). At this point the construction of the antenna is already finished, it only needs to be tuned to resonance for proper operation.

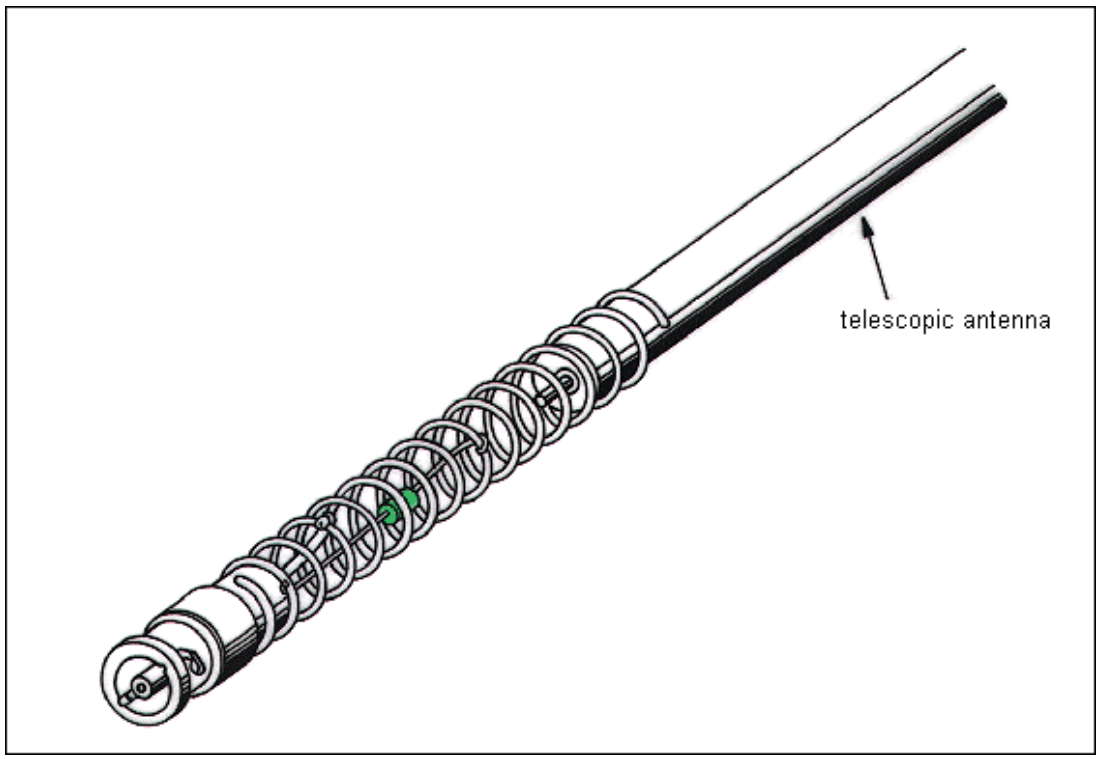


Figure 13.9 Final assembly of the antenna.

Tuning the antenna to resonance

Attach the antenna directly to the output connector of the SWR meter using the necessary adaptors. Similarly, connect the SWR meter to the transceiver using a short length of coaxial cable (see Figure 13.10).

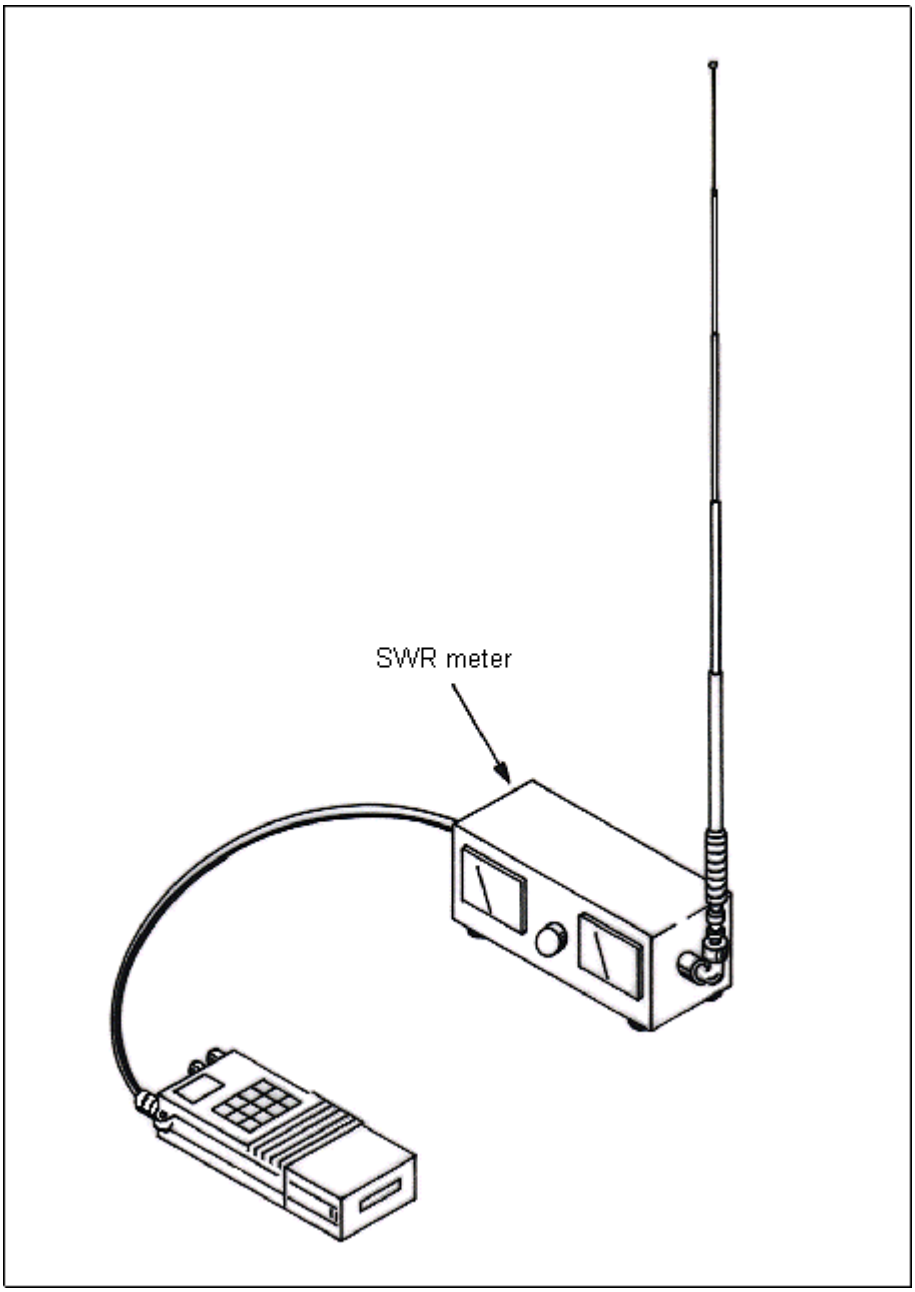


Figure 13.10 Preparing the PF-2C for resonance tuning.

Set the transceiver to the center frequency and key the PTT. Read the SWR response and note it on a chart similar to the one shown in Figure 13.11.

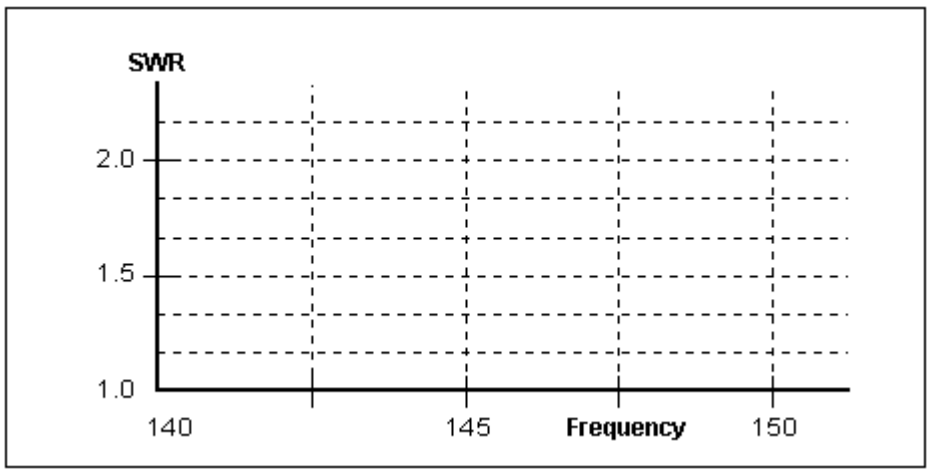


Figure 13.11 A sample chart for SWR readings.

Read all the SWR responses from the lowest frequency up to the highest frequency in the band and mark all the results on the chart until you get a response curve similar to the one shown in Figure 13.12.

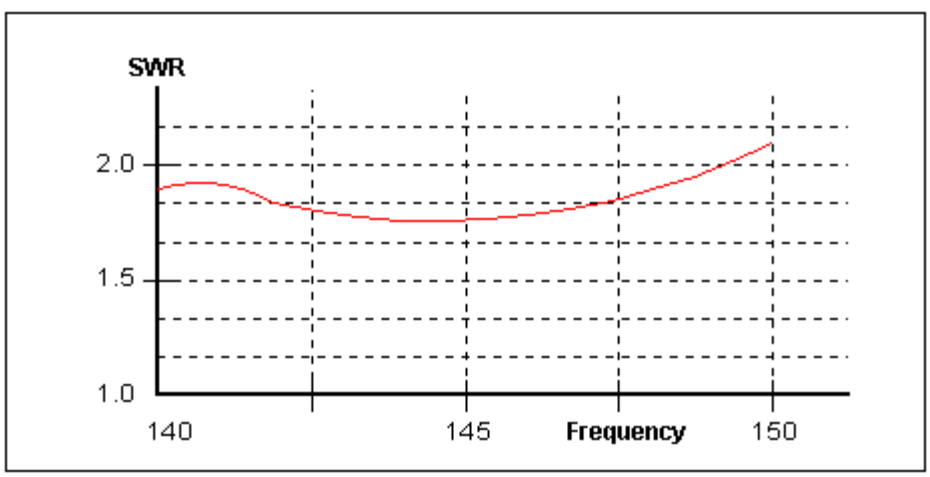


Figure 13.12 A sample of an SWR curve.

Resolder the capacitor's lead to a different point or tap either to the left or right of the original tap. If you have moved the tap to the right and the SWR went up then obviously you must move the tap to the left. Key again the PTT and mark the SWR responses once again on the chart. Move the tap about 1/8" farther at a time (see Figure 13.13).

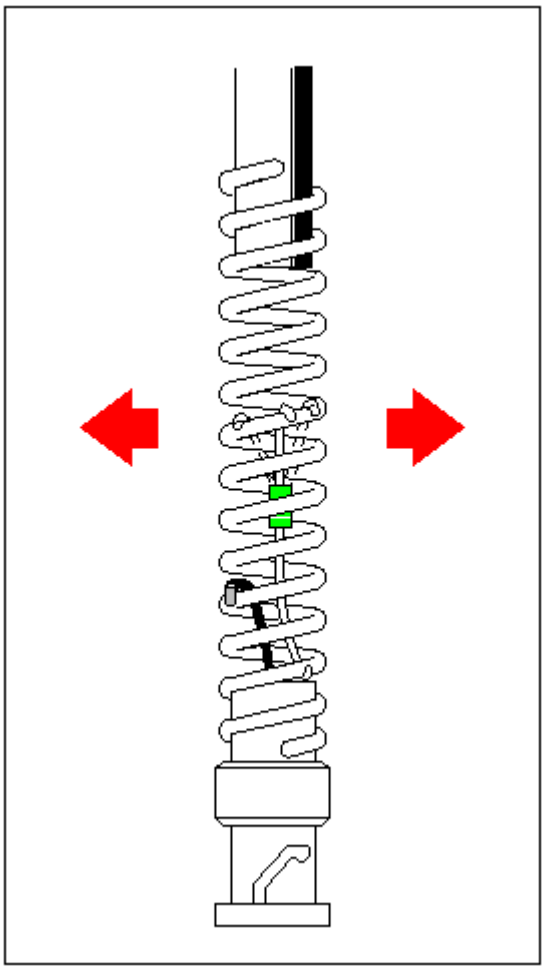


Figure 13.13 Resoldering the capacitor's lead to find the right tap.

Repeat the whole process until you find the point in the coil that results to a very low SWR reading on the center frequency and relatively balanced responses on the extreme ends of the band. If you have followed the instructions in constructing this antenna carefully, it is possible to get an SWR response of 1.1 at the center frequency and 1.5 at extreme ends of the band similar to the response curve shown in Figure 13.14.

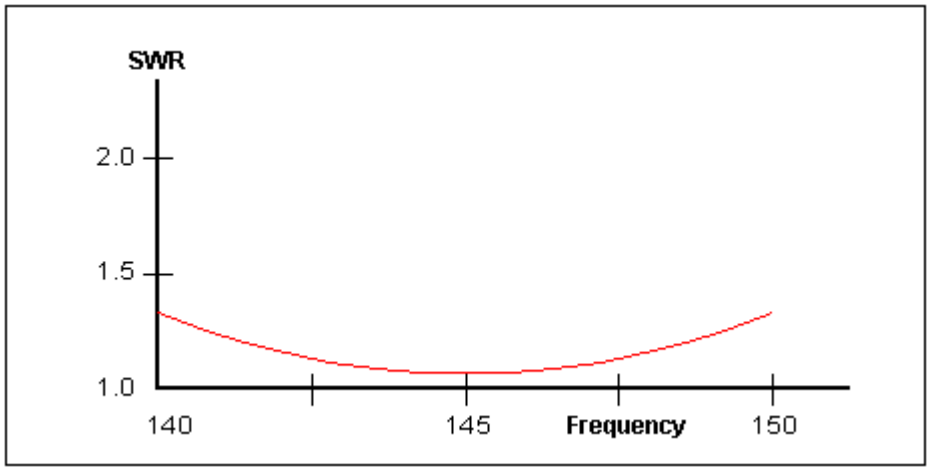


Figure 13.14 A sample of a good SWR response.

After you have found the right tap, solder it to the coil spring permanently. Insert the spring coil into the heat shrinkable tube and heat the tube over a flame or with a blow drier. Rotate the antenna and the tube continously while being heated to get an even shrinking of the tube (see Figure 13.15).

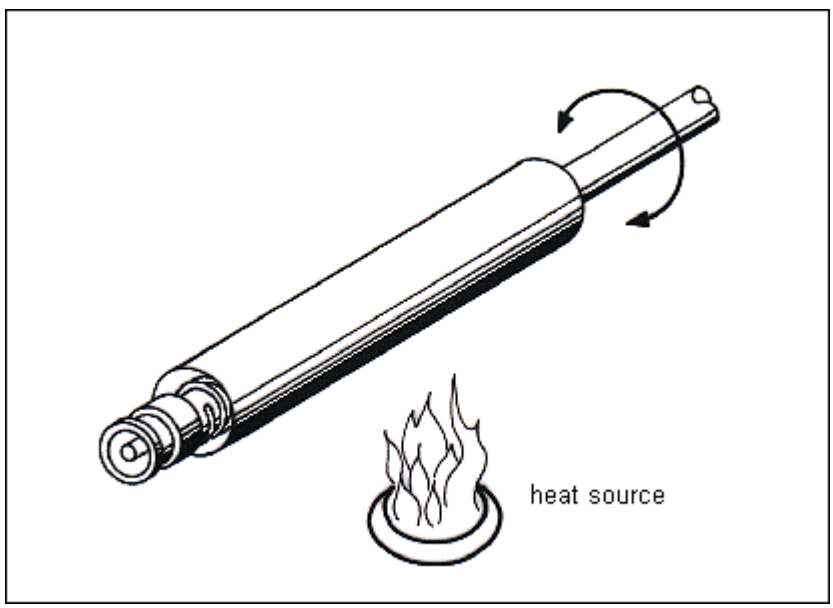
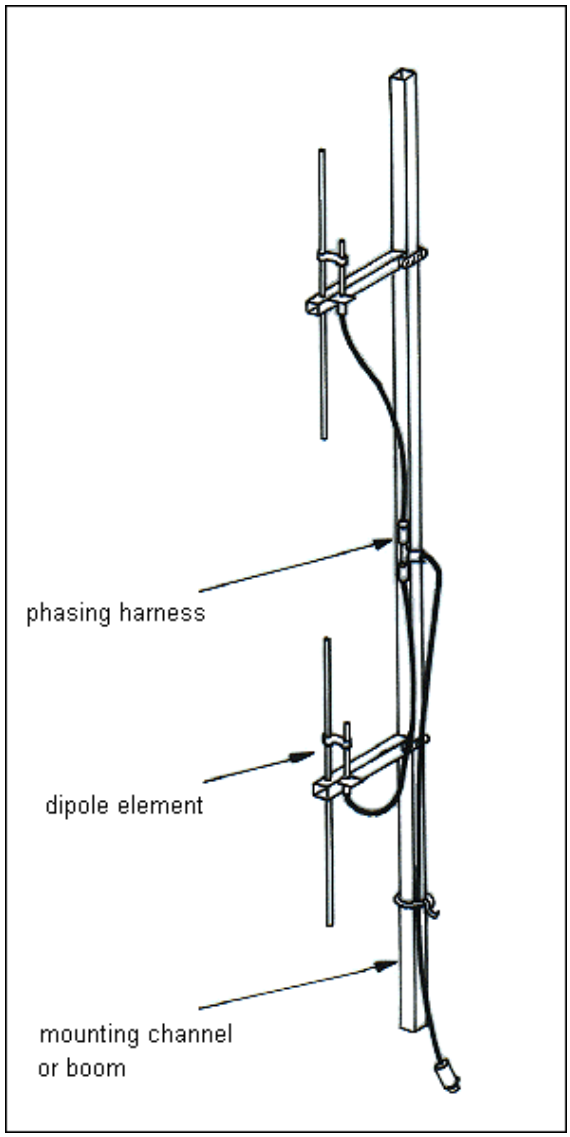


Figure 13.15 Heating the shrinkable tube.

14 COLLINEAR ANTENNA

Model SD-22 (2 stacked dipole)

A collinear antenna is made up of a multiple number of dipoles mounted in a common structure with their axis arranged in one straight line. The dipole elements are always driven in phase otherwise the array simply becomes a harmonic type antenna. A collinear array is a broadside radiator, meaning the direction of maximum radiation is at right angles to the line of the antenna.



When mounted vertically, it radiates an omni-directional pattern. One advantage of this design is its ability to attain high gain. When dipole elements are stacked collinearly, the power gain increases in direct proportion to the number of dipoles used. Obviously, this type of antenna is limited to fixed installation only because of its mechanical construction.

An actual working design of a collinear array is presented here. It has two identical dipoles fed with a coaxial phasing line or 'harness'. Each dipole element is tuned by a gamma matching system similar to that described in chapter 6. In fact, it is the same design of dipole just 'doubled' and fed simultaneously. This configuration gives a gain of 3 dB compared to a single dipole.

This model is dimensioned to operate in the frequencies of 140-150 MHz band. It has an SWR response of less than 1.5:1 over the entire band. Tuning procedure is similar to that described for dipole model DP-2.

Materials needed

Most of the materials needed to build the antenna model SD-22 are the same with those needed for antenna model DP-2 except for the mounting channel. The mounting channel for SD-22 is shorter being only 8 inches long and has slits on two sides instead of two holes.

Additionally, another long square channel is needed to mount the two dipole elements into a single mast. Also, a system of phasing harness made of coax cable is required to feed the two dipoles simultaneously. In short the additional materials needed for SD-22 are as follows:

Quantity	Description	Dimensions
2 pcs.	Square aluminum channel	1" x 1" x 8"
1 pc.	Square aluminum channel	1" x 1" x 115"
2 pcs.	Hose clamp	2 - 1/2" clamping capacity
6 pcs.	BNC VHF male connector	
2 pcs.	BNC 'T' connector	

Construction

Follow the procedures for constructing the dipole antenna model DP-2 and make two identical dipoles. The mounting channel for model SD-22 is slightly different and is described in the following illustration (see Figure 14.2).

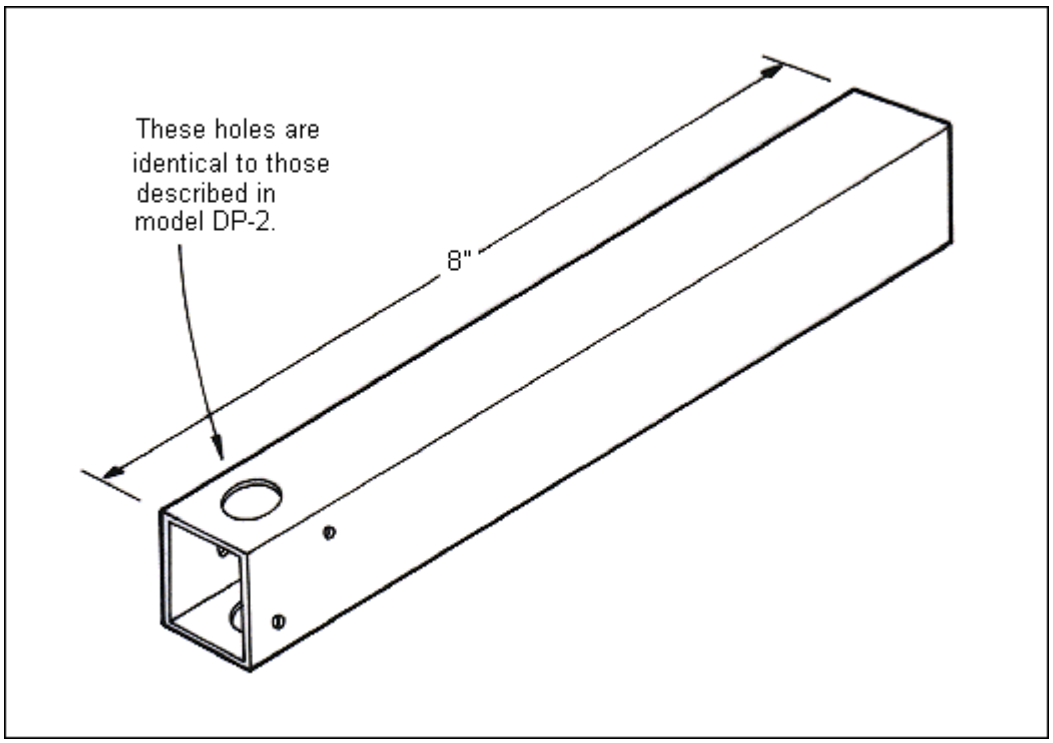


Figure 14.2 Mounting channel dimensions.

Saw shallow slits at two sides of the channel using a hacksaw. A hose clamp will be inserted into these slits for the purpose of mounting the channel to the supporting mast (see Figure 14.3).

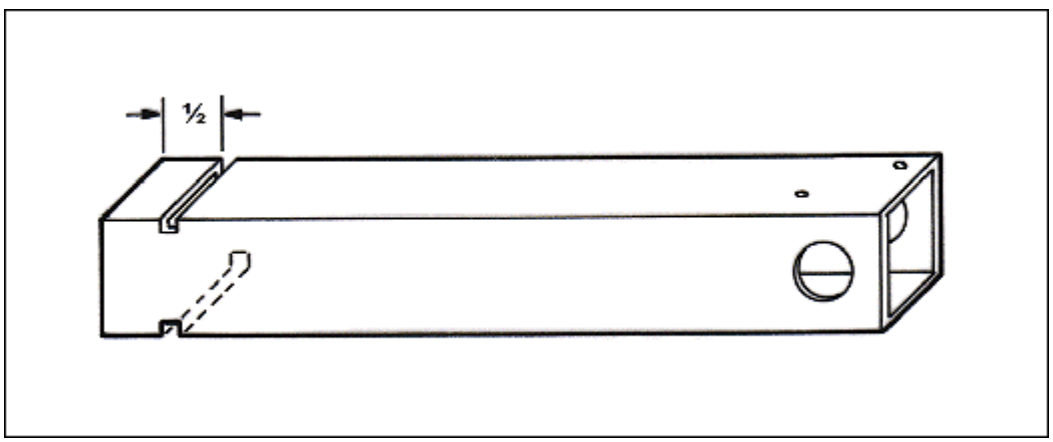


Figure 14.3 Saw slits at two sides of the channel.

Assemble the two dipole elements following the procedures described for antenna model DP-2. After the dipoles are completed, insert the hose clamps through the slits in the channel (see Figure 14.4).

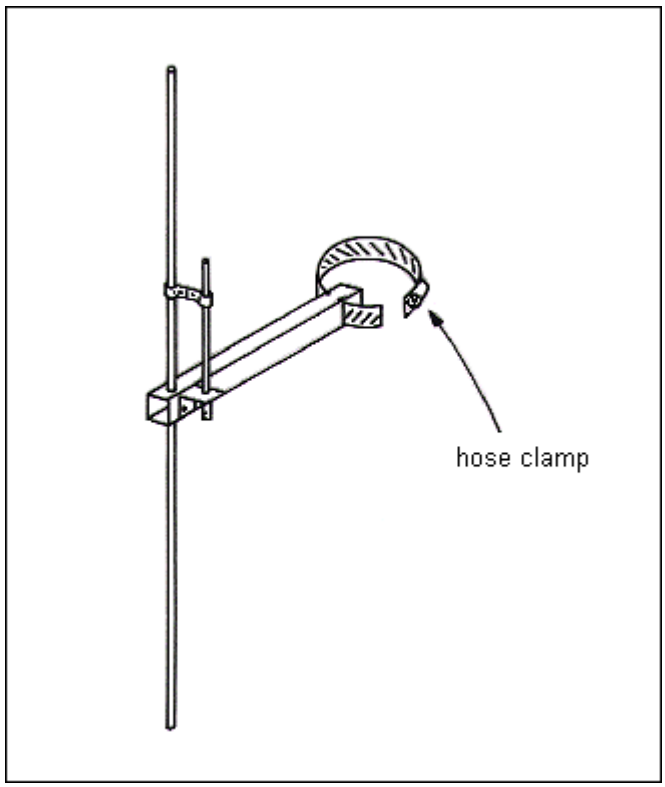


Figure 14.4 Assembled dipole element with hose clamp.

Mount the two dipoles to the aluminum supporting mast following the dimensions shown in Figure 14.5. Wrap the two hose clamps around the body of square channel mast and tighten the clamps to hold the dipole elements rigidly.

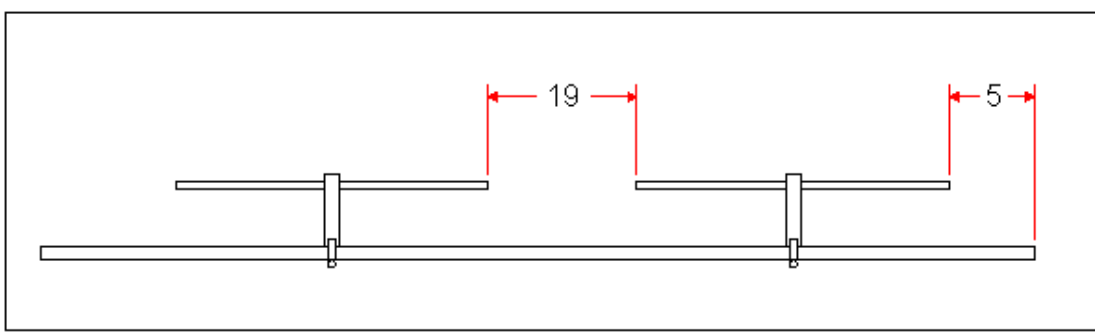


Figure 14.5 Mounting the dipoles to the aluminum mast.

Next, construct the phasing harness using RG-58/U coaxial cables and the appropriate connectors (see Figure 14.6).

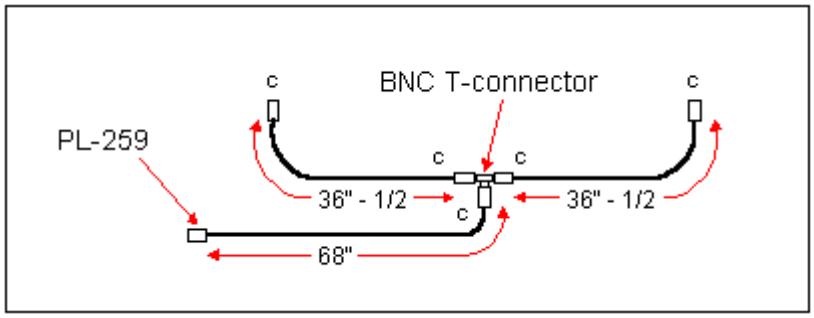


Figure 14.6 Constructing the phasing harness.

Finally connect the phasing harness to the two dipole elements and secure it to the support mast with plastic binders (see Figure 14.7).

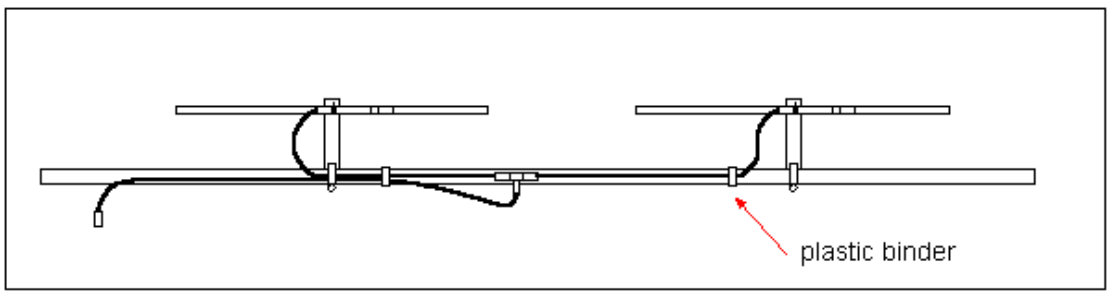


Figure 14.7 Connecting the phasing harness to the antenna.

Tuning the antenna

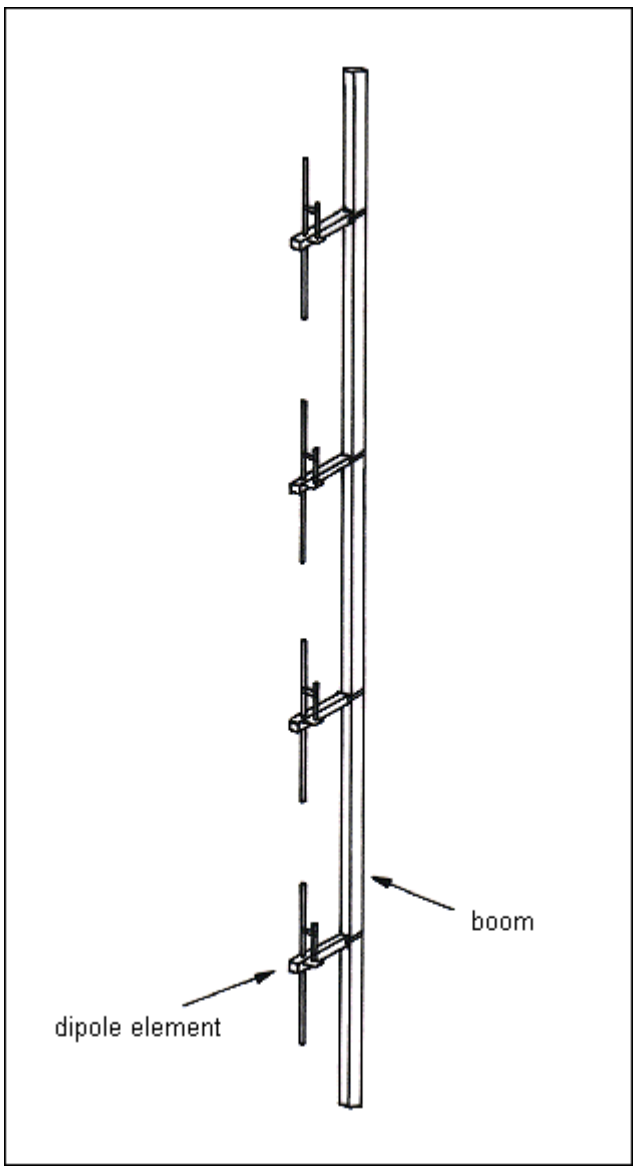
The tuning procedure for the antenna model SD-22 is similar to the procedure for tuning the antenna model DP-2. In tuning the SD-22 however, the two dipoles have to be tuned simultaneously. You have to do a lot of shuttling back and forth between the two dipoles before you can achieve a good match. If it is not practical to tune the antenna right in the main mast, then it can be tuned on the ground by placing it in a horizontal position with the dipoles facing upward. The antenna must be elevated to not less than 1 meter above the ground supported by non-metallic materials like wooden benches for example.

15 STACKED DIPOLE ARRAY

Model SD-24 (four stacked dipole)

This model demonstrates the capability of a simple dipole to attain high power gain by simply stacking identical units into a single structure and feeding them all simultaneously with a phasing harness. This arrangement is also called collinear array.

As stated earlier in chapter 14, power gain in a collinear array increases in direct proportion to the number of dipole elements used. However, in order to construct a practical phasing harness, the number of dipole elements installed cannot be simply dictated by personal choice. The correct method is to double the original number of dipole units - meaning, it the original array has two dipole elements installed, then the next array must have four dipoles and the next must have eight dipoles and so on. Everytime the number of dipole elements used is doubled, the power ratio is also doubled.



NOTE:
The power ratio is not numerically the same with the dB figure. For accurate computations refer to Appendix.

The particular four-element array presented here has a power gain of 6 dB. A collinear array having eight dipole elements will have a power gain of 9 dB. An array with elements in excess of eight is rarely constructed because of the inherent mechanical problems encountered in erecting structures of this size. Most collinear antennas are mounted vertically to effect an omni-directional pattern of radiation.

The model SD-24 is specifically dimensioned to operate in the frequencies of 140-150 MHz band. If properly tuned, this array exhibits an SWR of less then 1.5:1 over the entire band. The procedure for tuning this antenna to resonance is similar to the procedure for model SD-22.

Materials List

The necessary materials in building this antenna are the same with those needed for SD-22 being its extended version. The square channel used to mount the four dipoles is larger and twice longer than the one used for SD-22. Additional set of phasing harness is also needed to feed the four dipole elements simultaneously.

Additional materials for Model SD-24 are as follows:

1 pc.	**Square aluminum channel**	**1-1/2" x 1-1/2" x 235"**
4 pcs.	**BNC male connectors**	
2 pcs.	**BNC 'T' connectors**	

Construction

Construct the four dipoles following the procedures described for models DP-2 and SD-22. Mount the four dipoles to the aluminum supporting channel by using hose clamps. The antenna elements must be attached to the mast with the proper distances from each other (see Figure 15.2).

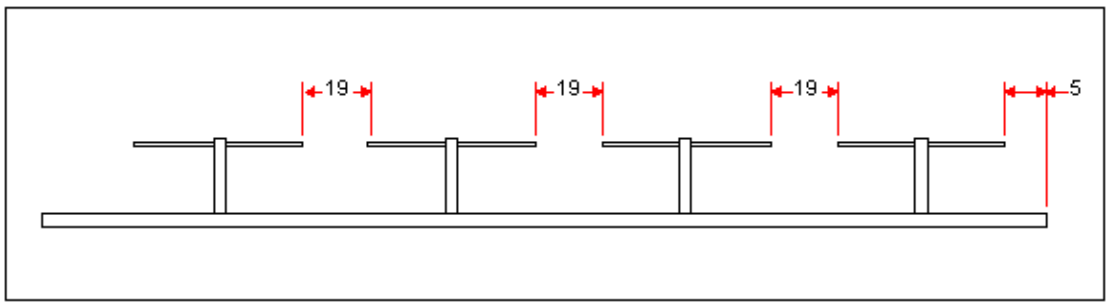


Figure 15.2 Mounting the four dipole elements on the mast.

Construct the phasing harness as shown and attach it to the four dipoles in similar fashion to model SD-22 antenna (see Figure 15.3).

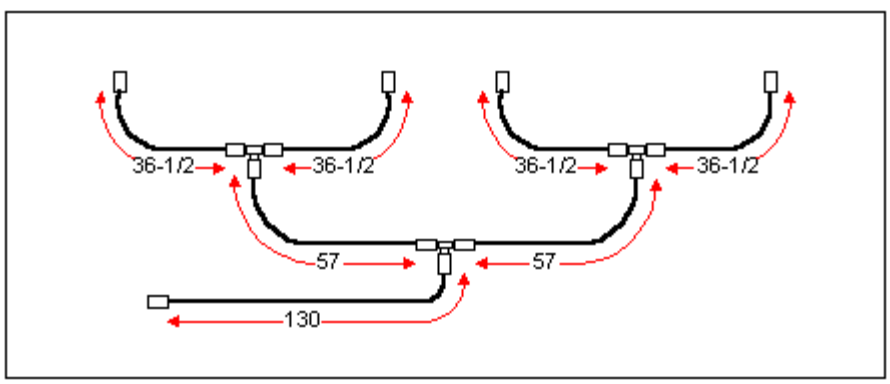


Figure 15.3 Constructing the phasing harness.

Tuning SD-24 to resonance

Tuning the antenna model SD-24 to get a good match is similar to procedures for tuning antenna model SD-22.

16 YAGI-UDA ANTENNA

Model YG-23 (3 element beam)

A yagi-uda antenna is a type of an array having one active dipole and two or more parasitic elements. It was named after the two Japanese physicists who invented it. The basic yagi is one of the highest gain antennas yet developed. Several factors affect the performance of a Yagi. Among these are the number of elements used, their diameter, and the spacing between them.

A basic hafwave dipole is cut to resonance at the center of the frequency band and is utilized as the driven element. High gain is attained by the addition of parasitic elements positioned either in front or behind the driven element. These parasitic elements are called directors and reflectors depending on their length and positioning with respect to the drive element. The reflector is longer by approximately 5 % and is positioned behind the driven element. The director on the other hand is cut shorter by approximately 5% and is positioned at the front of the driven element. The combination of these elements produce the directivity of the radiated signal thus resulting to higher power gain. However, the radiation pattern becomes uni-directional and the much desired omni-pattern is completely lost.

Maximum radiation of signal is now concentrated at the front of the antenna and there is only minimum radiation at the back. The ratio between the radiated signal at the front and the radiated signal behind it is called 'front to back ratio'. Radiation is weakest at the sides of the yagi and these points are called 'null points'. The ratio between the radiated signal at the front and the radiated signal at the sides is called 'front to side ratio'.

This highly directive and uni-directional characteristics of a yagi antenna necessitates the use of a rotator device in order to beam it to the direction of the station in contact. If a rotator device is not used, the high gain character of a yagi becomes useless unless the antenna is intended to be permanently beamed to a single direction such as in the case of fixed point-to-point communication.

The dimension of model YG-23 is specially cut to resonate in the frequency band of 140-150 MHz. If properly tuned, it exhibits an SWR of less than 1.5:1 over the entire band. It has a gain of approximately 7.3 dB compared to a standard dipole reference. This is only a basic configuration of a yagi and its gain and directivity can be increased by adding more directors at the front. Detailed information for the exact dimensions of additional director elements and their spacing is given in chapter 17. These yagi dimensions are based on the information published by the National Bureau of Standards.

Scaled down Appearance

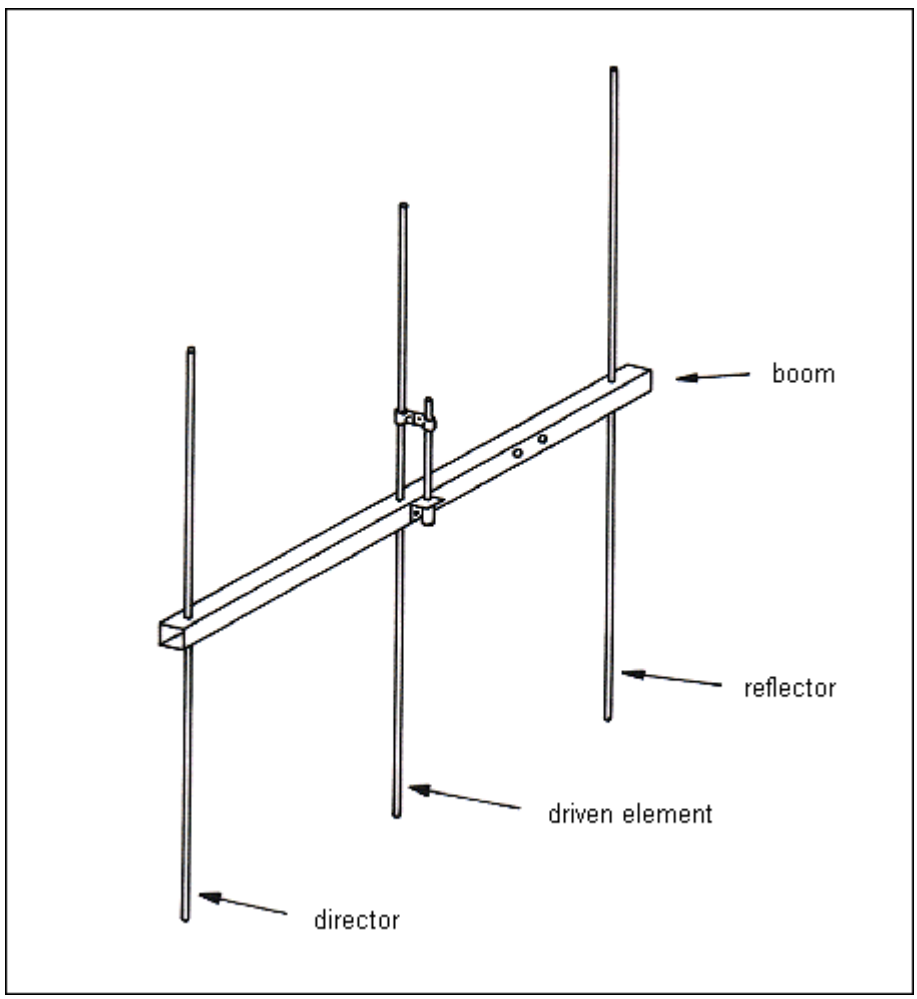


Figure 16.1 Yagi antenna model YG-23

Materials List

Quantity	Specification/Description	Dimensions
1	Aluminum tube 3/8" od*	3 feet 4"
1	Aluminum tube 3/8" od*	3 feet 2-3/16"
1	Aluminum tube 3/8" od*	3 feet 7/8"
1	Aluminum square channel	1" x 1" x 2 feet and 32-1/4"

Other materials used in constructing the antenna DP-2 are also needed for this yagi antenna except for the mounting channel.

* od - outside diameter

Construction

Cut the three tubes to their exact lengths and drill a hole (1/8" diameter) thru and thru at its middle length. The shortest tube will be used as a **director** element, the longest tube will the **reflector** element and the medium length will be the **driven** element (see Figure 16.2).

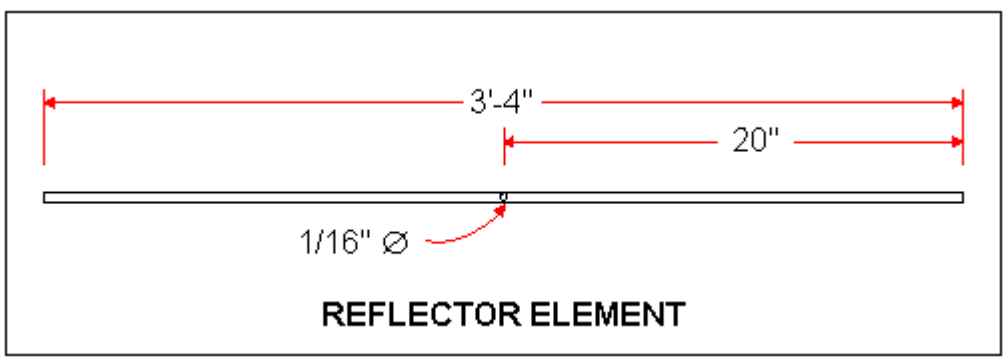


Figure 16.2 Preparing the reflector element.

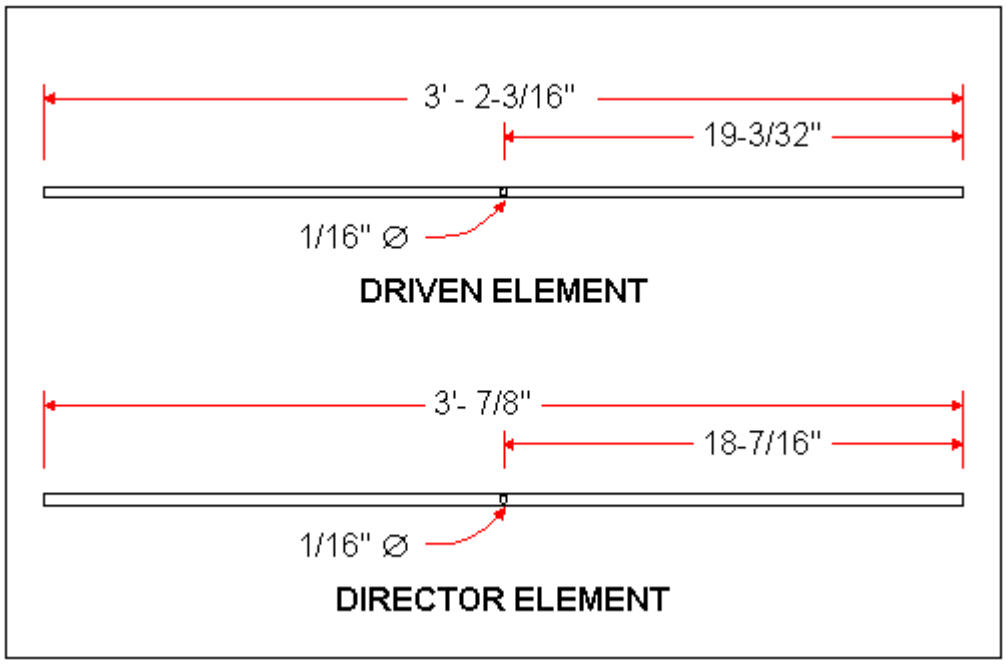


Figure 16.3 Preparing the driven and director elements.

Cut the aluminum mounting tube or boom to 2 feet and 11 inches long and drill three holes thru and thru at one side. The holes must have a diameter of 3/8" or enough to accomodate the diameter of the tube that will be inserted into it. Follow the dimensions shown. Drill also three 3/8" diameter holes at the same point where the larger holes are but at one side. The axis of the smaller holes must cross the axis of the larger holes (see Figure 16.4).

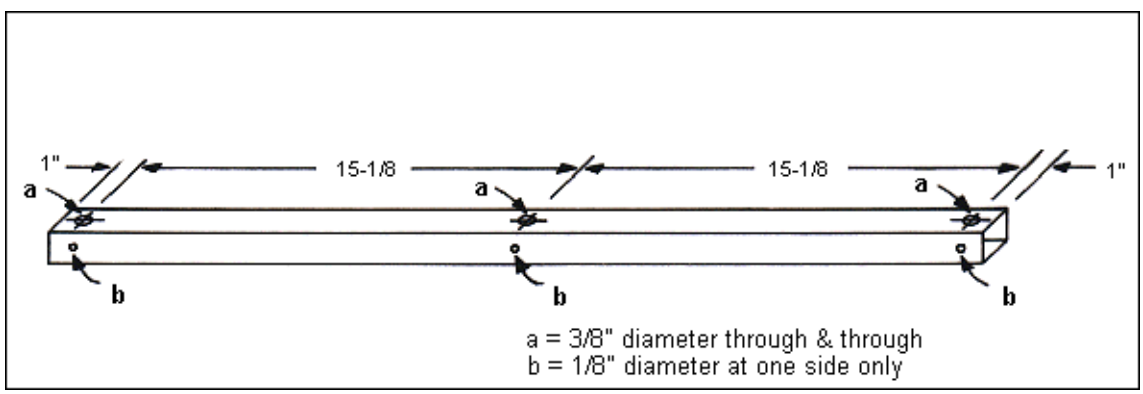


Figure 16.4 Preparing the boom.

Insert the aluminum tubes to the boom following the illustration for the proper arrangement of the elements. Secure the tube to the boom by placing the screws through the holes at the sides similar to the method of attaching the dipole element of the antenna model DP-2 (see Figure 16.5).

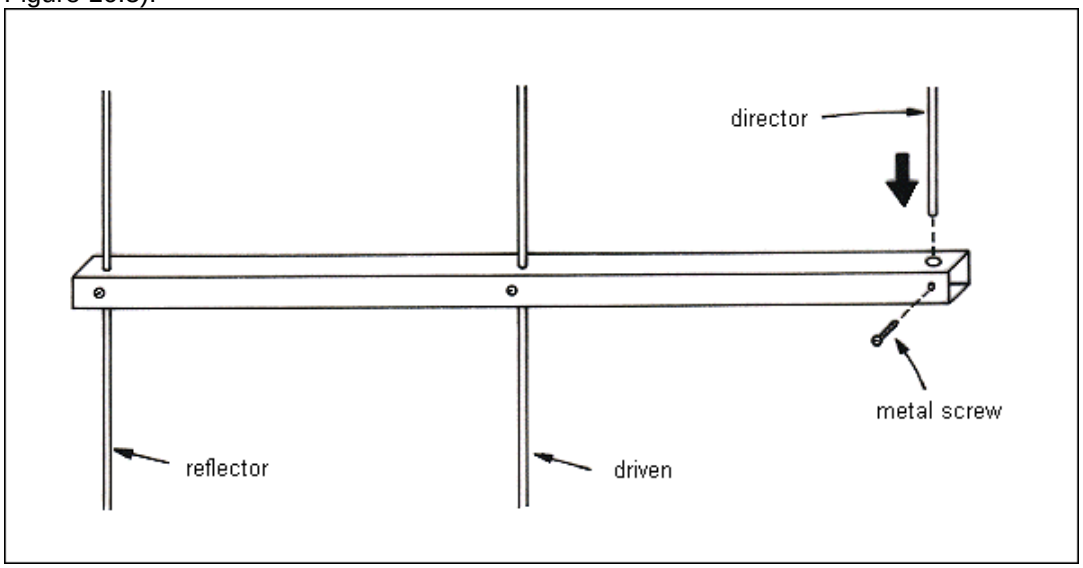


Figure 16.5 Assembling the antenna elements to the boom.

Complete the attachments of the **driven** element following the procedures described for model DP-2. All other materials and dimensions (e.g. gamma, bracket, connector, clamp, etc.) are similar to those used for the dipole elements of DP-2 (see Figure 16.6).

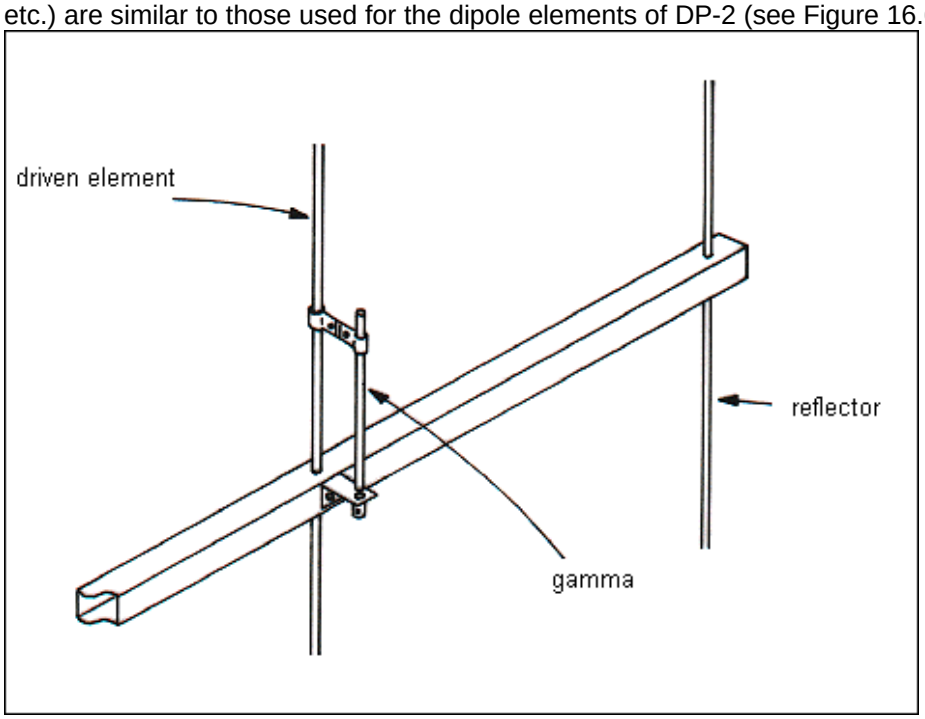


Figure 16.6 Complete assembly of the driven element.

Installation of YG-23

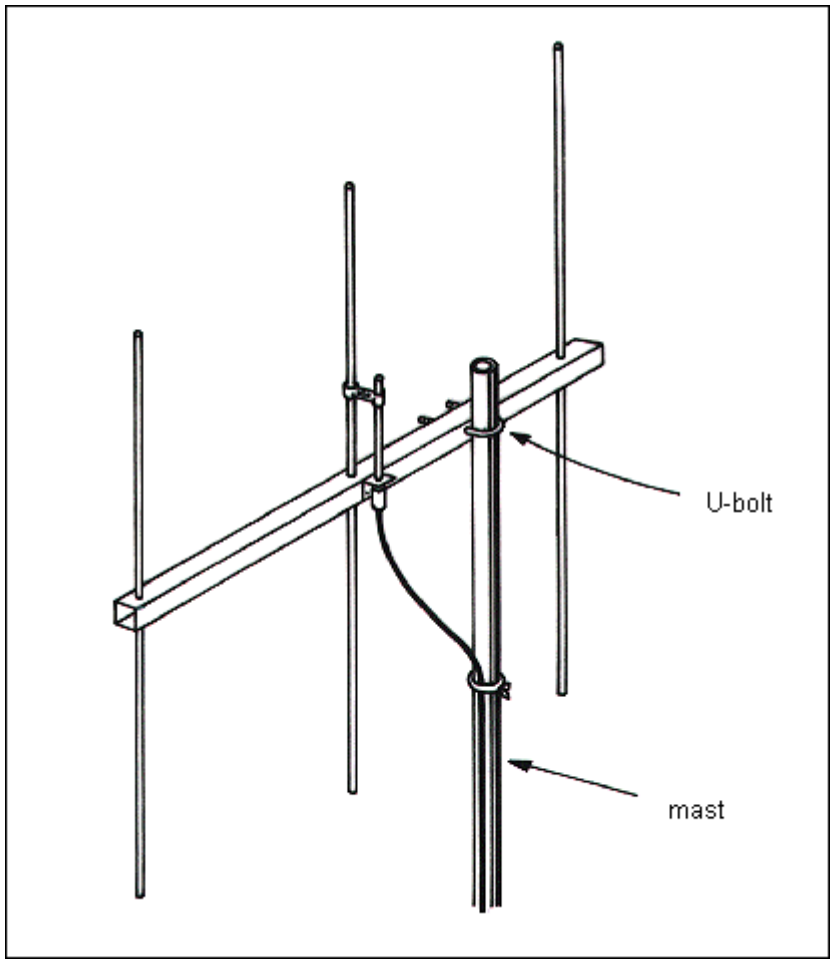


Figure 16.7 Installing the YG-23 to the mast.

Tuning YG-23 to resonance

The tuning procedure for the antenna model YG-23 is similar to the procedure for tuning antennas model DP-2, SD-22 or SD-24. The most important instruction to keep in mind is to tune the antenna while directly attached to the mast where it will be permanently installed whenever practical.

17 MULTI-ELEMENT YAGI-UDA ANTENNA ARRAY

The Yagi-Uda antenna or simply 'yagi' model YG-23 described in the preceeding chapter gives a fairly high gain figure in a very compact and easy to construct antenna. By adding more director elements at the front and extending the boom length of the yagi, you can achieve a very high gain figure from this type of antenna.

The following table shows the exact element lengths and dimensions for various yagi antennas based on the NBS standard. Two sets of dimensions are given, one set for the type of yagi antenna with elements that are insulated from the boom and another set for the yagi antenna with elements directly attached to the metal boom. The latter is widely popular among the antenna constructors because it is easier to construct and eliminates the need for individual insulators.

The construction of the reflector, driven element, gamma match, director elements, assembly of the whole antenna and tuning procedures are basically the same with the model YG-23.

NBS YAGI STANDARD DIMENSIONS

Boom Length	Boom Diameter	Element Diameter	Insulated Elements	Reflector	Driven	Dir.1	Dir.2	Dir.3	Dir.4	Dir.5
5'5-9/16" (0.8λ)	1"	3/8" od*	YES	3'4"	3'2-3/16"	3"7/8"	3'11/16"	3' 7/8"	-	-
			NO	3'4-5/8"	-do-	3'1-1/2"	3'1-3/8"	3'1-1/2"	-	-
8'2-5/16" (1.2λ)	1"	3/8" od*	YES	3'4"	-do-	3' 7/8"	3'7/16"	3'7/16"	3'7/8"	-
			NO	3'4-5/8"	-do-	3'1-1/2"	3'1-1/8"	3'1-1/8"	3'1-1/8"	-
15'1/4" (2.2λ)	1-1/2"	3/8" od*	YES	3'4"	-do-	3'1-1/8"	3'5/16"	2'11-13/16"	2'11-1/4"	2'10-9/16"
21'10-1/16" (3.2λ)	1-1/2"	3/8" od*	YES	3'4"	-do-	3'7/8"	3'9/16"	2'11-3/4"	2'11-1/8"	2'10-7/8"
			NO	3'5-1/16"	-do-	3'1-15/16"	3'1-3/8"	3'13/16"	3'3/16"	3'
28'8-1/8" (4.2λ)	1-1/2"	3/8" od*	YES	3'3-3/8"	-do-	3'9/16"	3'9/16"	3'3/8"	2'11-5/8"	2'11-1/2"
			NO	3'4-1/2"	-do-	3'1-5/8"	3'1-5/8"	3'1-7/16"	3'11/16"	3'9/16"

Dir. 6	Dir. 7	Dir. 8	Dir. 9	Dir. 10	Dir. 11	Dir. 12	Dir. 13	Dir. 14	Dir. 15
-	-	-	-	-	-	-	-	-	-
-	-	-	-	-	-	-	-	-	-
-	-	-	-	-	-	-	-	-	-
-	-	-	-	-	-	-	-	-	-
2'10-9/16"	2'10-9/16"	2'10-9/16"	2'11-1/4"	2'11-13/16"	-	-	-	-	-
2'11-3/8"	2'11-3/8"	2'11-3/8"	3'	3'5/8"	-	-	-	-	-
2'10-9/16"	2'10-5/16"	2'10-5/16"	2'10-5/16"	2'10-5/16"	2'10-5/16"	2'10-5/16"	2'10-5/16"	2'10-5/16"	2'10-5/16"
2'11-5/8"	2'11-3/8"	2'11-3/8"	2'11-3/8"	2'11-3/8"	2'11-3/8"	2'11-3/8"	2'11-3/8"	2'11-3/8"	2'11-3/8"
2'11-1/8"	2'10-13/16"	2'10-9/16"	2'10-9/16"	2'10-9/16"	2'10-9/16"	2'10-9/16"	2'10-9/16"	-	-
3'3/16"	2'11-7/8"	2'11-5/8"	2'11-5/8"	2'11-5/8"	2'11-5/8"	2'11-5/8"	2'11-5/8"	-	-

* od - outside diameter

ELEMENT SPACING WITH RESPECT TO BOOM LENGTH

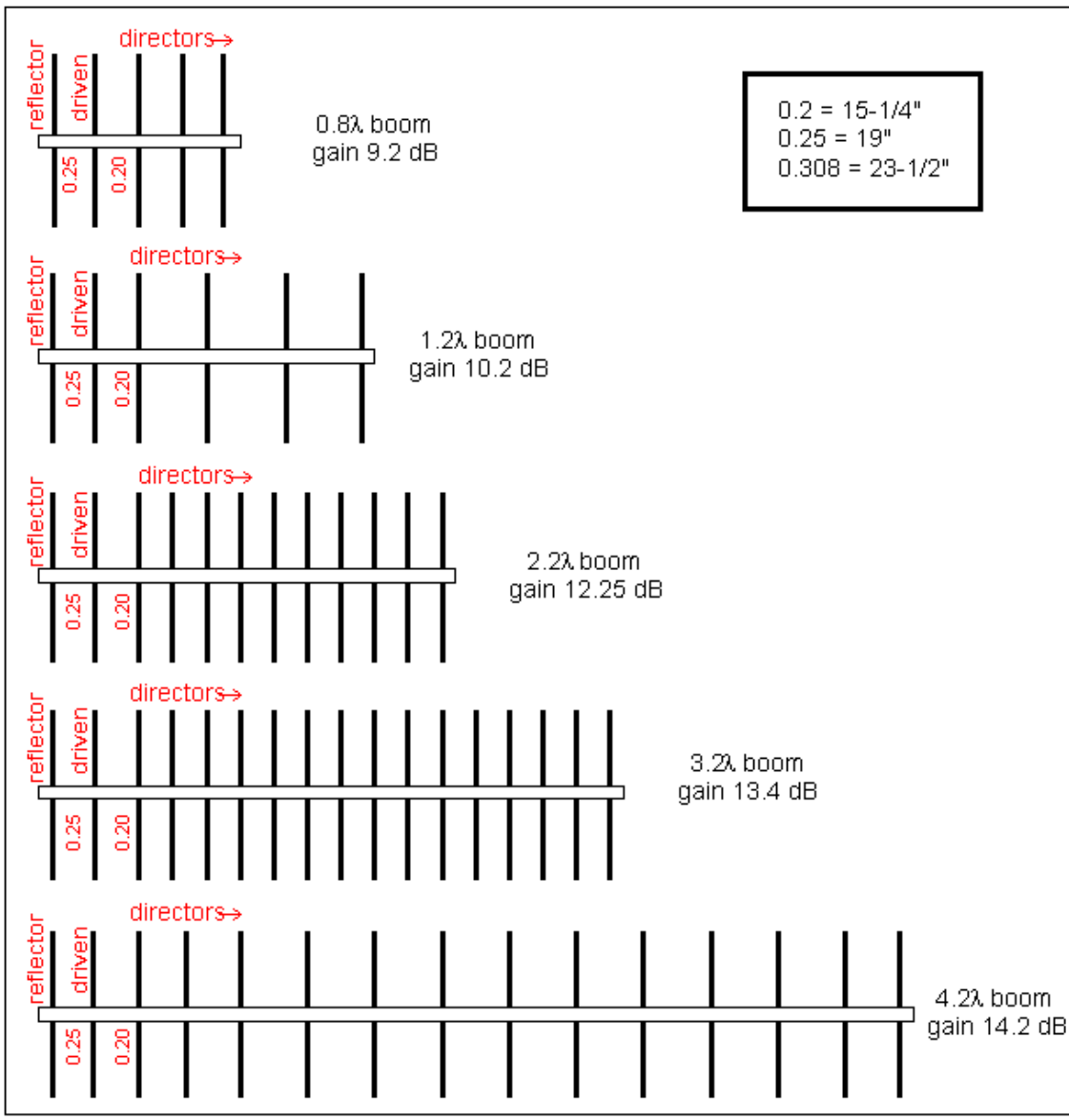


Figure 17.1 Element spacing of a Yagi antenna with respect to its boom length.

18 STACKING YAGI ANTENNAS

Stacking yagi antennas means multiplying the number of yagi antennas and feeding them all simultaneously. If a number of the yagi antennas is doubled, it will add an additional 3 dB to the original gain figure of the yagi. For example, if you feed two identical 3-element yagi which has a gain of 7.3 dB, it will give you a total gain figure of 10.3 dB. Similarly, a 17-element yagi with a gain of 13.4 dB will give a whooping 19.4 dB if stacked to four identical pieces!

In stacking yagis, the spacing between the antennas is important. The distance between two yagis stacked side by side must not be less than 1 wavelength or it must be approximately 77 inches. The distance between the tips of the elements in vertically stacked yagis must not be less than one-half wavelength or it must be approximately 38 inches (see Figures 18.1 and 18.2).

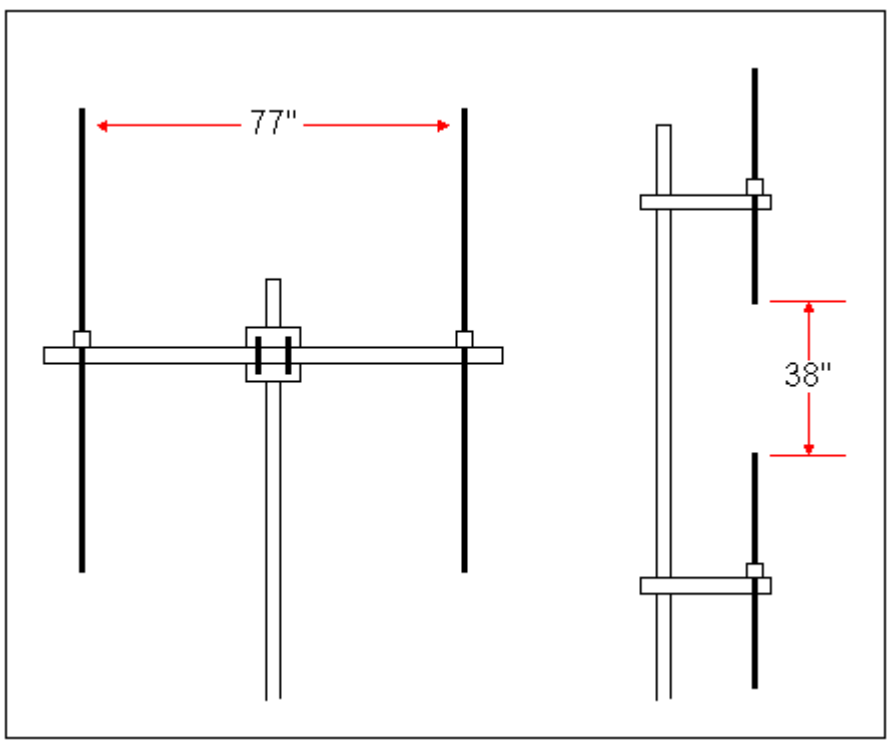


Figure 18.1 2 stacked yagis viewed at their boom ends.

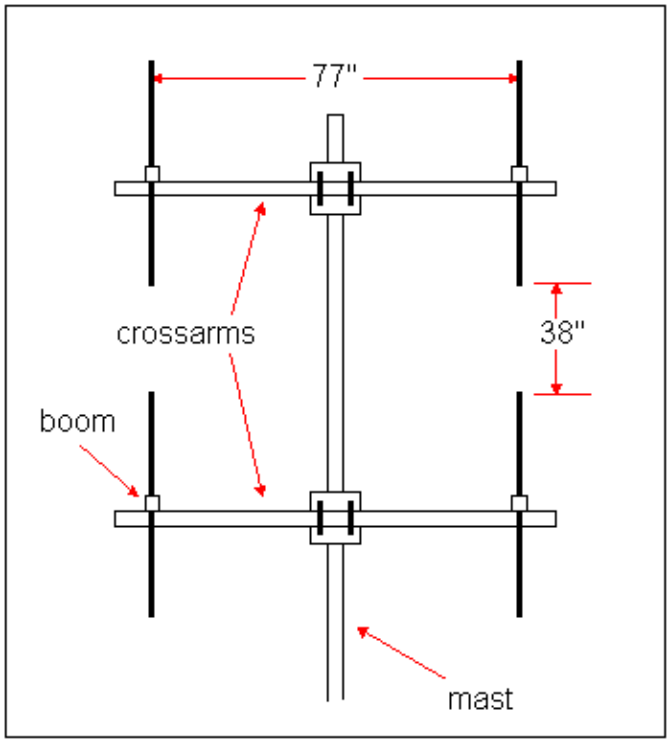


Figure 18.2 4 stacked yagis viewed at their boom ends.

All the yagis must be fed in phase with a phasing harness. For example, the configuration of a phasing harness for the 4 stacked yagis shown above is described in the following illustration.

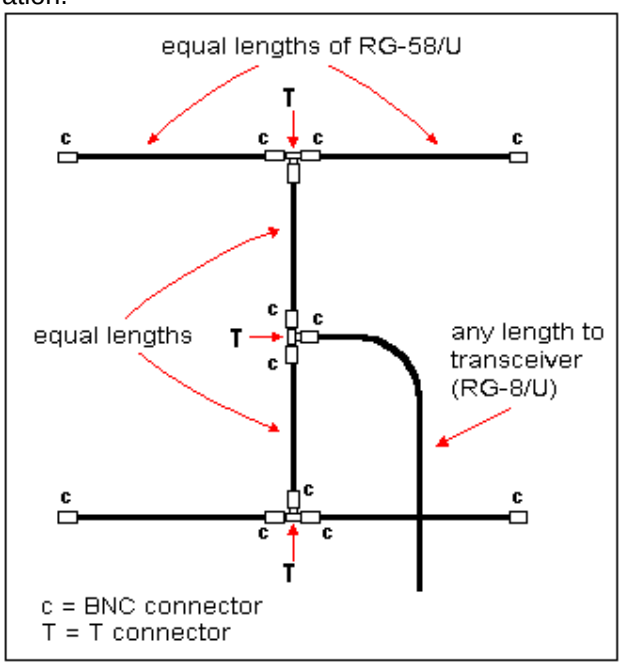


Figure 18.3 Phasing harness for a 4-stacked yagi antenna.

Mechanical Construction

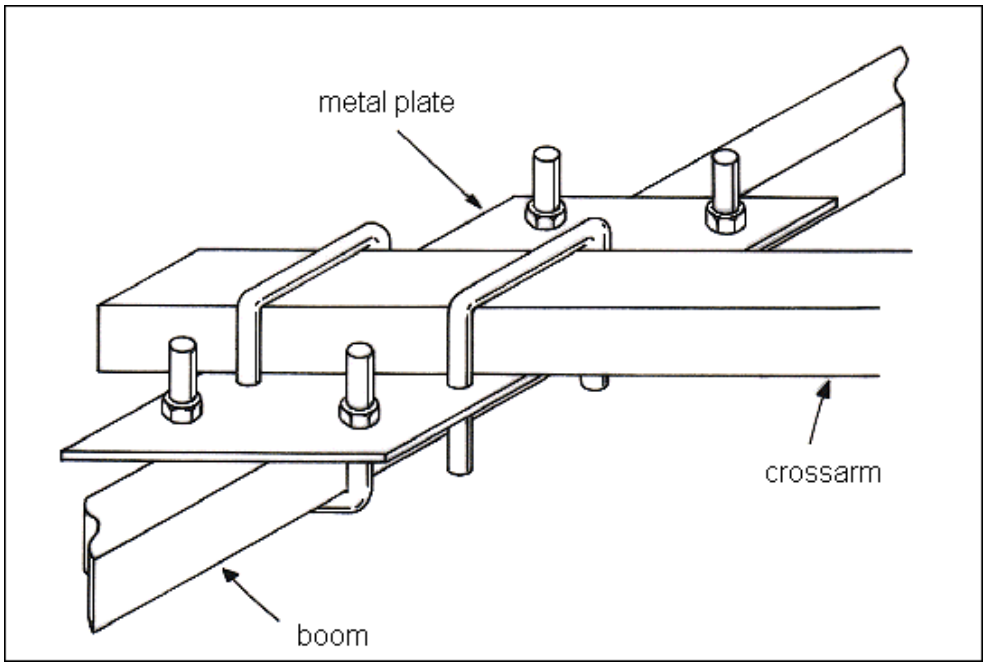


Figure 18.4 Attachment of the boom to cross-arm.

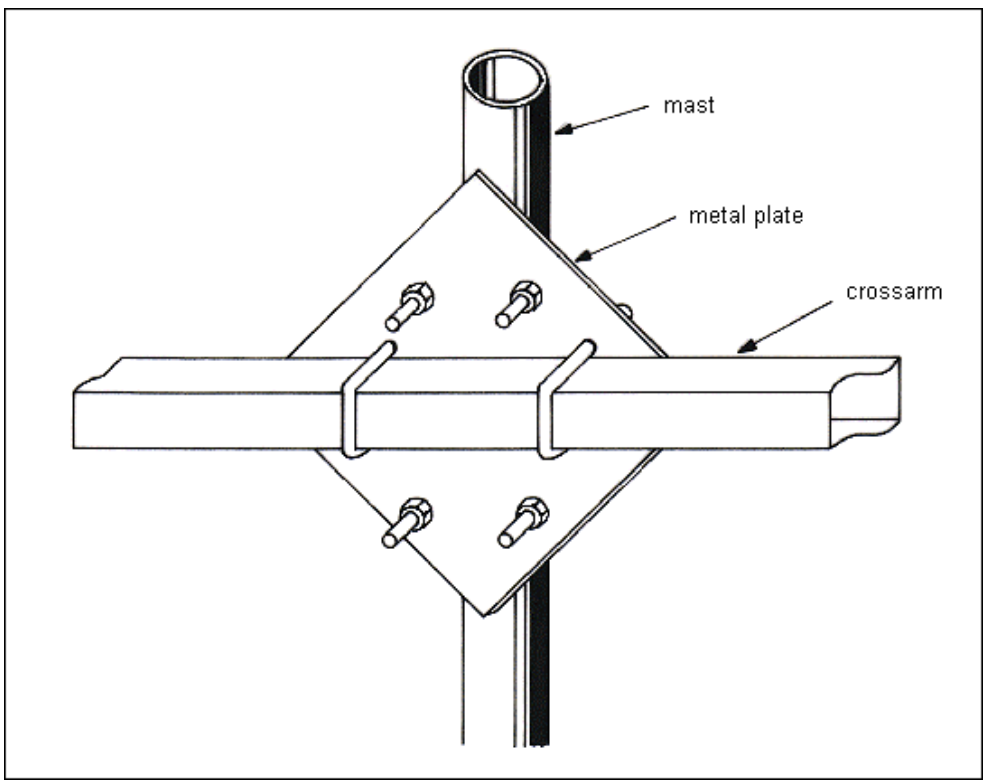


Figure 18.5 Attachment of the cross-arm to mast.

19 FORMULAS FOR CONVERTING ANTENNA DESIGNS FOR OTHER FREQUENCY BANDS

The dimensions of the antenna elements are generally derived from the antenna's electrical wavelength. **The electical wavelength of a certain frequency is slightly different from its wavelength in free space**, where the former is the wavelength of the signal present in the **physical** conductor of the antenna and is somewhat shorter.

The formula to get the electrical wavelength of a frequency is:

In feet:

$$\frac{936}{F_C \text{ (MHz)}} = \lambda \text{ (feet)}$$

In meters:

$$\frac{286}{F_C \text{ (MHz)}} = \lambda \text{ (meters)}$$

- F_C is the center frequency of the band expressed in Megahertz.
- Lambda λ is the symbol for the wavelength expressed either in feet or meters depending on the particular units used.

■ For example, the wavelength of 145 MHz is:

$$\frac{936}{145} = 6.46 \text{ feet or } 77.52 \text{ inches}$$

Other symbols of wavelength:

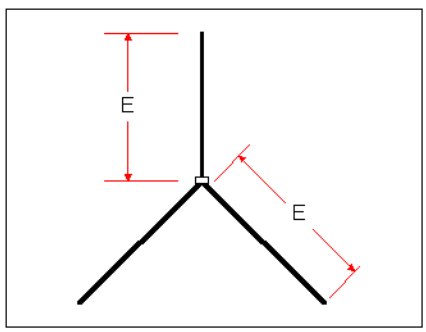
$\lambda/2$ or $(0.5)\lambda$ = halfwavelength
$\lambda/4$ or $(0.25)\lambda$ = quarterwavelength

Groundplane elements

The formula to get the length of each element of a groundplane antenna is:

$$\mathbf{E}\ (\text{feet}) = \frac{468}{Fc} \qquad \text{or} \qquad \mathbf{E}\ (\text{feet}) = 0.25\ \lambda$$

- To convert **E** (feet) to inches multiply it with 12.



- **Example:** Find the length of one groundplane element intended for 220 MHz.

Solution:

$$\mathbf{E}\ (\text{feet}) = \frac{468}{220} \qquad \textit{substituting the value of frequency}$$

$$\mathbf{E}\ (\text{feet}) = \frac{2.127}{2}$$

$$\mathbf{E}\ (\text{feet}) = 1.06 \qquad \textit{this is the length of the element expressed in feet}$$

To convert the result to inches:

$$\mathbf{E}\ (\text{feet}) \times 12 = \mathbf{E}\ (\text{inches})$$

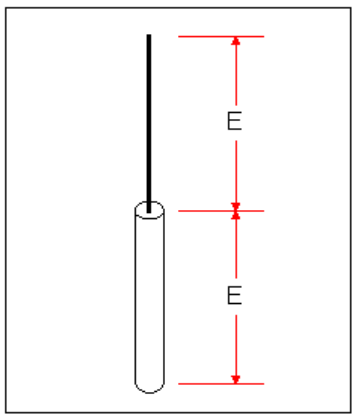
$$1.06 \times 12 = 12.72$$

$$\mathbf{E} = \mathbf{12.72\ inches} \qquad \textit{this is the length of the element expressed in inches}$$

Coaxial dipole elements

Formulas to find the elements of a coaxial dipole:

$$\mathbf{E}\ (\text{feet}) = \frac{468}{Fc} \qquad \text{or} \qquad \mathbf{E}\ (\text{feet}) = 0.25\ \lambda$$



- **Example:** Find the length of the element for a coaxial dipole intended for 110 MHz (also known as aircraft band).

Solution:

$$\mathbf{E}\ (\text{feet}) = \frac{468}{Fc}$$

$$\mathbf{E}\ (\text{feet}) = \frac{468}{110}$$

substituting the value of frequency

$$\mathbf{E} = 2.125 \text{ feet}$$

this is the length of the coaxial element expressed in feet

Convert the result to inches:

$$\mathbf{E}\ (\text{feet}) \times 12 = \mathbf{E}\ (\text{inches})$$

$$2.125 \times 12 = 25.5$$

$$\mathbf{E} = \mathbf{25.5\ inches}$$

this is the final value expressed in inches

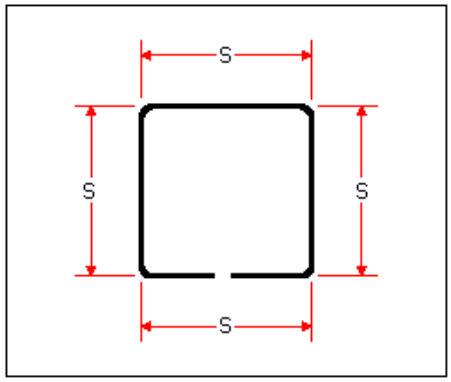
Quad loop antenna element

The formula to get the length of each side of the loop element for the Quad loop antenna:

$$\mathbf{S}\ (\text{feet}) = \frac{486}{Fc}$$

or

$$\mathbf{S}\ (\text{feet}) = \frac{0.25\ \lambda}{2}$$



■ **Example:** Find the length of one side of the quad loop intended for 155 MHz (VHF commercial band).

Solution:

$$\mathbf{S}\ (\text{feet}) = \frac{468}{Fc}$$

$$\mathbf{S}\ (\text{feet}) = \frac{468}{155} \qquad \textit{substituting the value of frequency}$$

$$\mathbf{S}\ (\text{feet}) = \frac{3.02}{2}$$

$$\mathbf{S} = 1.51 \text{ feet} \qquad \textit{this is the length of one side of the loop element expressed in feet}$$

Convert the result to inches:

S (feet) x 12 = **S** (inches)

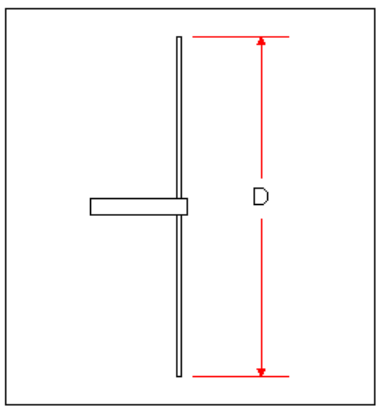
1.51 x 12 = 18.12

S = **18.12 inches** *this is the final value expressed in inches*

Dipole element

The formula to get the length of the dipole element for the antennas DP-22, SD-22, and SD-24:

$$\textbf{D}\ (\text{feet}) = \frac{486}{Fc} \qquad \text{or} \qquad \textbf{D}\ (\text{feet}) = 0.5\ \lambda$$



■ **Example:** Find the length of the dipole element intended for 195 MHz.

Solution:

$$\textbf{D}\ (\text{feet}) = \frac{468}{Fc}$$

$$\textbf{D}\ (\text{feet}) = \frac{468}{195}$$

D (feet) = 2.4 feet — *this is the length of the dipole element expressed in feet*

Convert the result to inches:

D (feet) x 12 = **D** (inches)

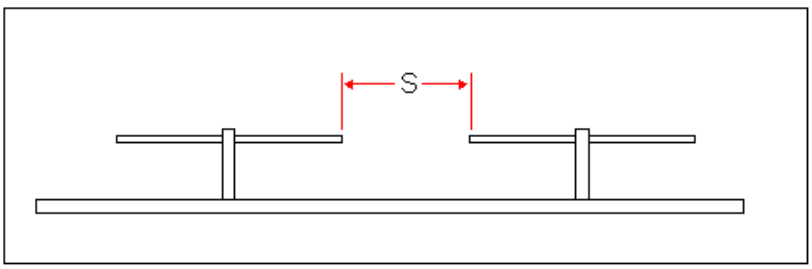
2.4 x 12 = 28.8

D = **28.8 inches** — *this is the final value expressed in inches*

Spacing between dipoles' ends

The formula to get the correct spacing between the ends of two dipoles in a collinear array such as SD-22 or SD-24:

$$\mathbf{S}\ (\text{feet}) = \frac{468}{Fc \quad 2} \qquad \text{or} \qquad \mathbf{S}\ (\text{feet}) = 0.25\ \lambda$$



■ **Example:** Find the spacing between two dipoles designed for 220 MHz.

Solution:

$$\mathbf{S}\ (\text{feet}) = \frac{468}{Fc \quad 2}$$

$$\mathbf{S}\ (\text{feet}) = \frac{468}{220 \quad 2}$$

$$\mathbf{S}\ (\text{feet}) = \frac{2.127}{2}$$

S = 1.053 feet *this is the length of the dipole element expressed in feet*

Convert the result to inches:

S (feet) x 12 = **S** (inches)

1.053 x 12 = 12.45

S = 12.45 inches *this is the final value expressed in inches*

NOTE:

This spacing is only for the minimum allowable between the ends of the dipole elements in a collinear antenna. Optimum spacing is within 0.25λ and 0.5λ.

Discone dimensions

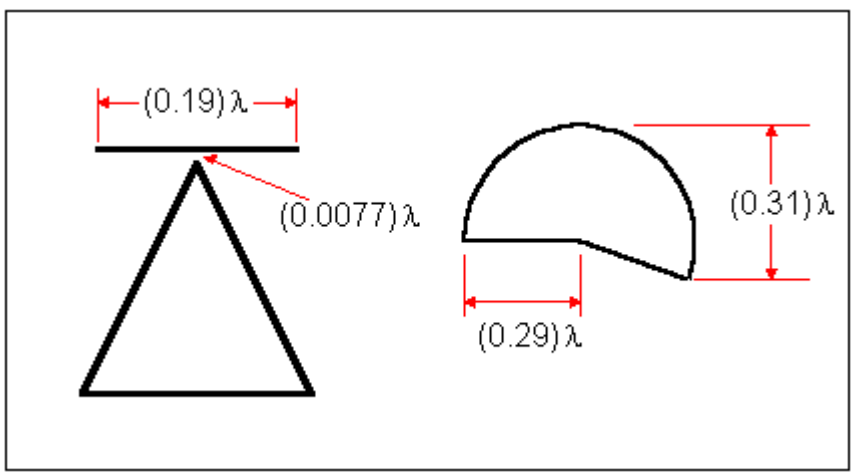
To get the exact lengths and dimensions of the discone antenna, first compute the wavelength of the lowest targeted frequency by using this formula:

$$\lambda \text{ (wavelength in feet)} = \frac{936}{Fc}$$

Formulas for dimensions:

The disc element diameter is **0.19** of the wavelength.
The cone element length is **0.29** of the wavelength.
The spacing between the disc element and the apex of the cone element is **0.0077** of the wavelength.

The diameter of the mounting tube is not critical.



■ **Example:** Find the dimensions of a discone antenna with a cut-off frequency of 120 MHz (upper portion of the aircraft band).

Solution:

First find the wavelength of the frequency by using the formula below and convert the result to inches.

$\frac{936}{Fc} = \lambda$ (wavelength in feet)

$\frac{936}{120} = 7.8$

7.8 feet = λ *this is the wavelength of the frequency in feet*

Convert it to inches:

7.8 x 12 = 83.6

λ (inch) = **83.6 inches** *wavelength of the frequency in inches*

From the result above you can proceed to compute the dimensions:

DISC ELEMENT DIAMETER:

Formula: 0.19 x λ (inch) = disc diameter

(substitute value of λ)

0.19 x 83.6 = 15.384

15.382 inches *diameter of the disc*

LENGTH OF CONE ELEMENT:

Formula: 0.29 x λ (inch) = length of cone element

(substitute value of λ)

0.29 x 83.6 = 24.44

24.44 inches *length of the cone element*

SPACING BETWEEN DISC AND CONE'S APEX:

Formula: 0.0077 x λ (inch) = spacing

(substitute value of λ)

0.0077 x 83.6 = 0.64

0.64 inches *spacing between disc and cone's apex*

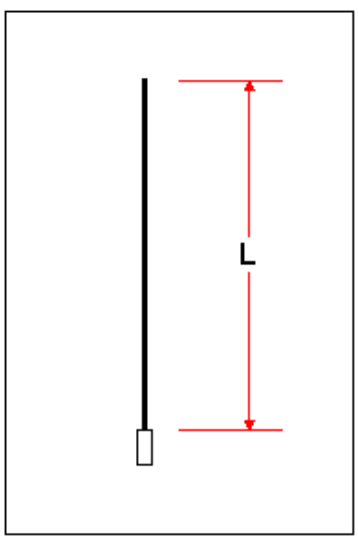
NOTE:

In calculating the dimensions of a discone, you must always use the value of the lowest frequency you intend to operate (cut-off frequency).

5/8 Wave radiator element

The formula to get the length of the radiator element of a 5/8 element wavelength antenna such as in WA-2 and PF-2C:

L (feet) = 0.65 x λ or **L** (feet) = $\frac{(5 \times \lambda)}{8}$



■ **Example:** Find the length of the radiator element intended for 160 MHz (VHF commercial band).

Solution:

First find the wavelength of the center frequency by using the formula below and convert the result to inches.

$$\frac{936}{Fc} = \lambda \text{ (wavelength in feet)}$$

$$\frac{936}{160} = 5.85$$

5.85 feet = λ *this is the wavelength of the frequency in feet*

Convert it to inches:

5.85 x 12 = 70.2

λ (inch) = **70.2 inches** *wavelength of the frequency in inches*

Finally, find the length of the 5/8 radiator element by using the following formula:

L (inch) = 0.65 x λ (inch)

L (inch) = 0.65 x 70.2

L = 45.6 inches *length of the radiator element*

APPENDIX

POWER RATIO TO DECIBEL CONVERSION

	Decimal Increments									
Ratio	0.0	0.1	0.2	0.3	0.4	0.5	0.6	0.7	0.8	0.9
1	0.00	0.41	0.79	1.14	1.46	1.76	2.04	2.30	2.55	2.79
2	3.01	3.22	3.42	3.62	3.80	3.98	4.15	4.31	4.47	4.62
3	4.77	4.91	5.05	5.19	5.32	5.44	5.56	5.68	5.80	5.91
4	6.02	6.13	6.23	6.34	6.44	6.53	6.63	6.72	6.81	6.90
5	6.99	7.08	7.16	7.24	7.32	7.40	7.48	7.56	7.63	7.71
6	7.78	7.85	7.92	7.99	8.06	8.13	8.20	8.26	8.33	8.39
7	8.45	8.51	8.57	8.63	8.69	8.75	8.81	8.86	8.92	8.98
8	9.03	9.08	9.14	9.19	9.24	9.29	9.34	9.40	9.44	9.49
10	10.00	10.04	10.09	10.13	10.17	10.21	10.25	10.29	10.33	10.37
x10	+10									
x100	+20									
x1000	+30									
x10,000	+40									
x100,000	+50									

How to use this chart:

The decibel value is read from the body of the table for the desired ratio, including the decimal increment. For example, a power ratio of 1.8 is equivalent to 2.55 dB. Value from the table may be extended as indicated at the lower left in each section. For example, a power ratio of 16 which is the same as 10 x 1.6 is equivalent to 10 + 2.04 = 12.04 dB.

Power loss in relation to the SWR figure in the transmission line:

SWR	Power lost in %
1 :1	0 %
1,3 :1	2 %
1,5 :1	3 %
1,7 :1	6 %
2 :1	11 %
3 :1	25 %
4 :1	38 %
5 :1	48 %
6 :1	55 %
10 :1	70 %

METRIC EQUIVALENTS

Most of the antenna dimensions described in this book are in English units. If the constructor wants to use the metric system, he can convert all the dimensions by using the following conversion guide:

English to Metric

Inch = 25.4 millimeters
Inch = 2.54 centimeters
Foot = 0.305 meter
Yard = 0.914 meter

Metric to English

Centimeter = 0.3937 inches
Meter = 39.37 inches
Meter = 3.28 feet
Meter = 1.094 yards

GLOSSARY OF ANTENNA TERMS

Actual ground The point within the earth's surface where effective ground conductivity exists. The depth of this point varies with the frequency, the condition of the soil and the geographical region.

Antenna An electrical conductor or array of conductors that radiates signal energy (transmitting) or collects signal energy (receiving).

Apex The feedpoint region of a discone antenna.

Apex angle The enclosed angle in degrees inside the cone element of a discone antenna and similar antennas.

Bandwidth The group of frequencies where the antenna functions efficiently.

Band A group of frequencies.

Coaxial cable Any of the coaxial transmission lines that has the outer shield (either solid or braided) in the same axis as the inner or center conductor. The insulating material can be air, helium, or solid dielectric compounds.

Collinear array A linear array of radiating elements (usually dipoles) with their axis arranged in a straight line. Popular in VHF and higher frequencies.

Conductor A metal body such as tubing, rod or wires that permits current to travel continously along its length.

Counterpoise A wire or group of wires mounted close to ground, but insulated from ground, to form a low impedance, high capacitance path to ground. Commonly used at medium frequency and high frequency to provide an effective ground for an antenna.

Dielectrics Various insulating materials used in antenna systems, such as found in insulators and transmission lines.

Dipole An antenna that is split exactly at the middle for connection to a feedline. Usually a halfwavelength in dimension. Also called a doublet.

Directivity The property of an antenna that concentrates the radiated energy to form one or more major lobes.

Director A conductor placed in front of a driven element to cause directivity. Frequently used singly or in multiples with Yagi or cubical quad beam antennas.

Direct ray Transmitted signal energy that arrives at the receiving antenna directly rather than being reflected from the ionosphere, ground or man made reflector.

Doublet see Dipole

Driven array An array of antenna elements which are all driven or excited by means of a transmission line.

Driven element The radiator element of an antenna system. The element to which the transmission line is connected.

Efficiency The ratio of useful output power to input power, determined in antenna systems by losses in the system, including in nearby objects.

Feeders Transmission lines of assorted type that are used to route RF power from a transmitter to an antenna, or from an antenna to a receiver.

Feedline see Feeders

Front to back The ratio of radiated power off the front to the back of a directive antenna. A dipole would have a ratio of 1 for example.

Front to side The ratio of radiated power between the major lobe and the null side of a directive antenna.

Gain Increase in effective radiated power in the desired direction of the major lobe.

Gamma match A matching system used with driven antenna elements to effect a match between the transmission line and the feed-point of the antenna. It consists of an adjustable arm that is mounted close to the driven element and in parallel with it near the feedpoint.

Groundplane A man made system of conductors placed below an antenna to serve as an earth ground.

Groundscreen A wire mesh groundplane.

Impedance The ohmic value of an antenna feedpoint, matching section or a transmission line. An impedance may contain reactance as well as resistance components.

Lambda Greek symbol for L used to represent a wavelength with reference to electrical dimensions in antenna work.

Line loss The power lost in a transmission line, usually expressed in decibels.

Line of sight Transmission path of a wave that travels directly from the transmitting antenna to the receiving antenna.

Load The electrical entity to which the power is delivered. The antenna is a load for a transmitter. A dummy load is a nonradiating substitute for an antenna.

Loading The process of transferring power from its source to a load. The effect of a load has on a power source.

Lobe A defined field of energy that radiates from a directive antenna.

Matching The process of effecting an impedance match between two electrical circuits of unlike impedance. One example is matching a transmission line to the feedpoint of an antenna. Maximum power transfer to the load (antenna system) will occur when a matched condition exists.

Null A condition during which an electrical property is at minimum. The null in an antenna radiation pattern is that point in the 360 degree pattern where minimum field intensity is observed. An impedance bridge is said to be 'nulled' when it has been brought into balance.

Parasitic array A directive antenna that has a driven element and independent directors or reflectors or both. The directors and reflectors are not connected to the feedline. A yagi antenna is one example. See also driven array.

Phasing lines Sections of transmission line that are used to ensure correct phase relationship between the bays of an array of antenna. Also used to effect impedance transformations while maintaining the desired array phase.

Quad Rectangular or diamond shaped fullwave loop antenna. Most often used with a parasitic loop director and a parasitic loop

reflector to provide approximately 8 dB of gain and good directivity. Often called the 'cubical quad'.

Radiation pattern The radiation characteristics of an antenna as a function of space coordinates. Normally, the pattern is measured in the far field region and is represented graphically.

Radiator A discrete conductor in an antenna system that radiates RF energy. The element to which the feedline is attached.

Reflector A parasitic antenna element or a metal assembly that is located behind the driven element to enhance forward directivity. Large man made structures may reflect radio signals.

Source The point of origination (transmitter or generator) of RF power supplied to an antenna system.

Stacking The process of placing similar directive antennas atop or beside one another forming a 'stacked array'.

SWR Standing wave ratio on a transmission line in an antenna system. More correctly, 'VSWR' or voltage standing wave ratio. The ratio of the forward to reflected voltage on the line and not the power ratio. A VSWR of 1:1 occurs when all parts of the antenna system are matched correctly to one another.

Velocity factor That which affects the speed of radio waves in accordance to the dielectric medium they are in. A factor of 1 is applied to the speed of light and radio waves in free space, but the velocity is reduced in various dielectric mediums such as transmission lines. When cutting a transmission line to a specific electrical wavelength, the velocity factor of the particular line must be taken into account.

VSWR Voltage standing wave ratio. See SWR.

Wave A disturbance that is a function of time or space or both.
A radio wave for example.

Wave front A continous surface that is the locus of points having the same phase at the same instant.

Yagi A directive, gain type of antenna that utilizes a number of parasitic directors and a reflector. Named after one of the inventors (Yagi and Uda).

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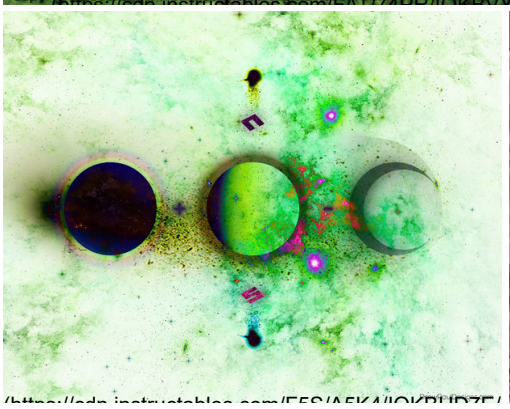

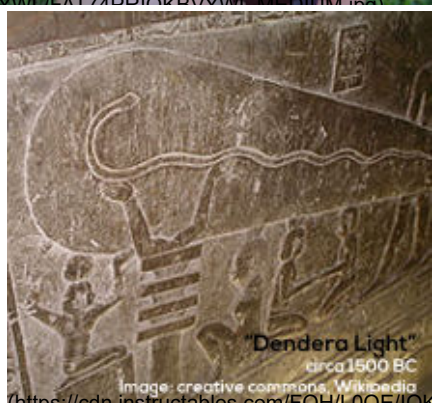

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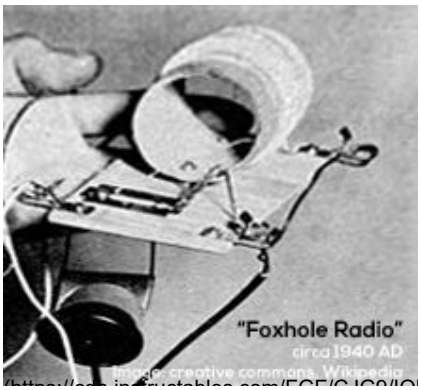


"Free energy from the air?, Yea, right!" Sardonic skepticism was my first reaction to this unusual concept, as well.

Though, its not so far out there, in fact. Light can be converted to DC current with solar panels, electricity can be converted to magnetism as I did in my last article (https://www.instructables.com/id/DIY-Electro-Magnetic-Levitation/), in a microphone sound waves are converted to an electrical signal (by vibrating a magnet near a coil (http://hyperphysics.phy-astr.gsu.edu/hbase/audio/mic.html)), solar rays can even be focused and converted to heat in awesome devices like this! (http://www.gosunstove.com/) When we think about it, energy is all around us and can be harvested in an enumerable many of ways.

Today, we are going to take a rather novel approach. We are going to build a device specifically designed to sense and capture a particular band of energy which is all around us.

The earth is magnetic and anyone who has ever used a compass knows this. Magnetic bodies in motion produce electricity, we can see this in any alternator, like the one in your car. So, therefore the earth is electric as well as magnetic, by definition.

Can we detect this energy? Yes, we sure can! Ever turn on a radio in the middle of nowhere and heard static? That is your radio picking up naturally occurring energy in the RF spectrum!

Can we use this energy to do work? Absolutely! This has been known for a long time. Crystal radios (https://en.wikipedia.org/wiki/Crystal_radio) have been around since before the 1930's and can run with no input energy other than the radio signal. Even when completely isolated, but from the atmosphere, a crystal radio will produce a voltage in the earpiece resulting in a sound (albeit and undesirable one).

Well, this is where it gets interesting...

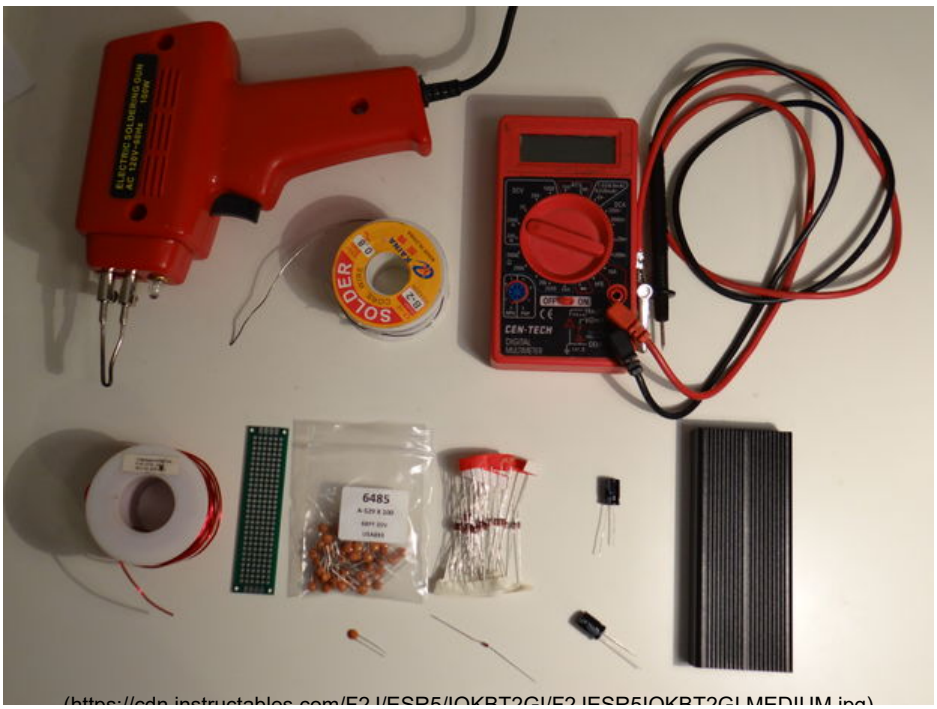
Can we replicate this effect? Yea, and with modern components like the high quality crystals found in germanium diodes, we can even increase efficiency. By applying this concept as a Crystal Energy Receiver we can take advantage of a wide range of energetic frequencies rather than tuning in to just one.

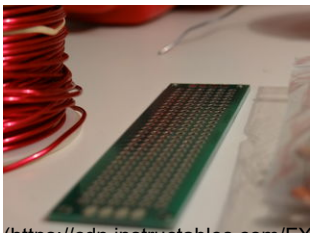
Can we scale it up? Definitely. Things like micro germanium diodes, high efficiency antennas and compact contemporary capacitors make the components that are required to build a crystal receiver fit in the palm of your hand. While there may or may not be a more efficient way, this renewable

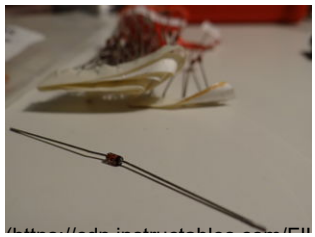
energy solution is simple to employ and can be scaled up, or down, indefinitely.

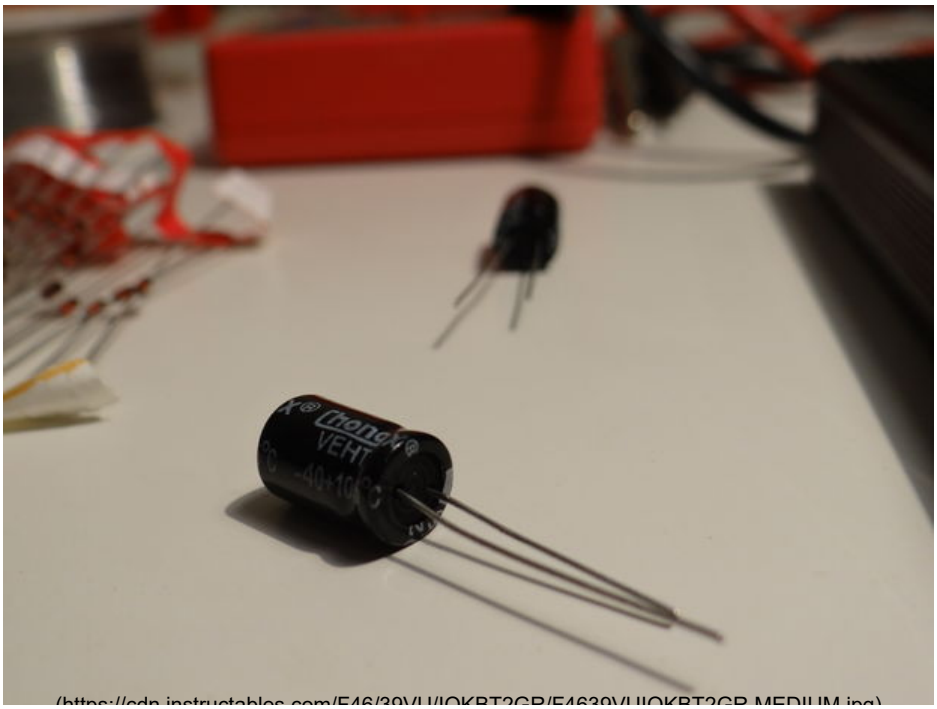
It sounds like we can build a Crystal Energy Receiver. Let's give it a shot...

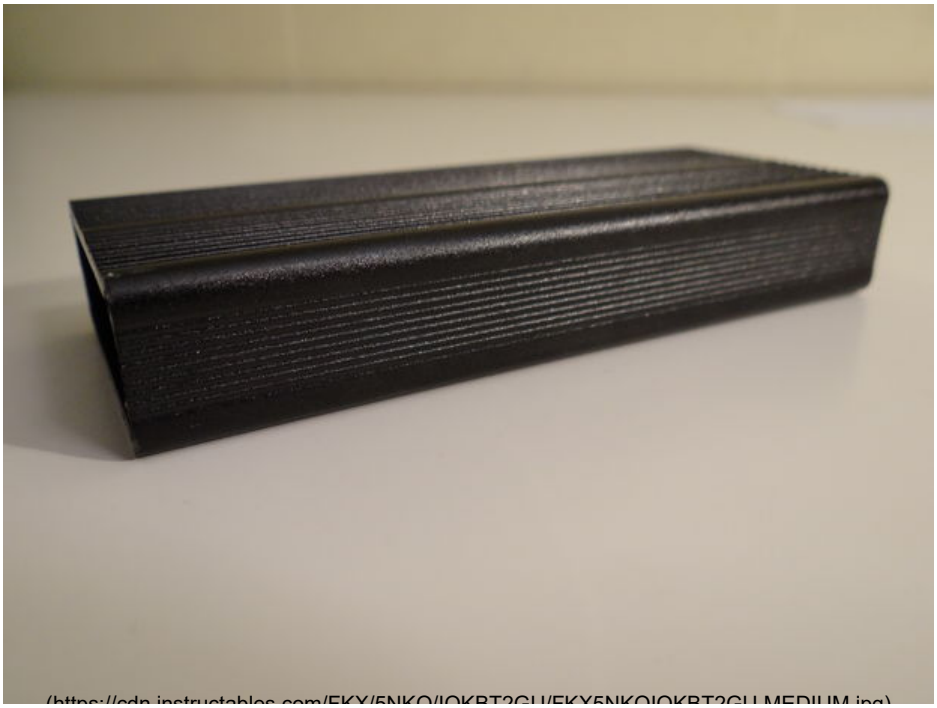
Step 1: What You'll Need











One of the reasons this particular renewable energy harvesting method is so viable is the relatively few and easy to obtain materials required.

The simplest crystal receiver design needs no power and can be built with only three parts: a coil, a crystal and a resistor. We're going to optimize that design in order to produce a cleaner and more reliable output signal by first polarizing the input amplitude, then rectifying and filtering the signal. Then we'll add an antenna, case and connections.

Get the circuit diagram here (http://www.drewpauldesigns.com/crystal-energy-receiver-kit.html)
Get the kit here (http://www.drewpauldesigns.com/crystal-energy-receiver-kit.html)

The parts for the circuit include:
(1) Circuit Board (http://www.drewpauldesigns.com/crystal-energy-receiver-kit.html)
(1) 10-18 gauge Copper Wire
(2-12+) Ceramic Capacitors (matched)
(2-6+) Electrolytic Capacitors (matched)
*note various types of capacitors can be used
(4) Germanium Crystal Diodes (1A+)

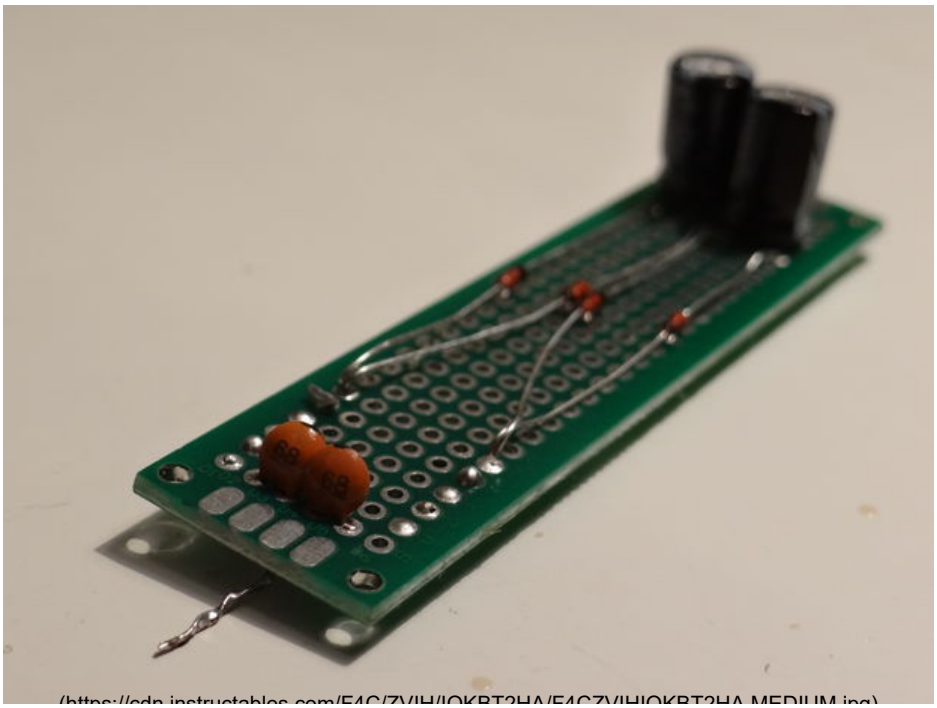
Total Unit Cost: +/- **$0.40** (USD, scaled for volume of 1,000+ units)

In addition, you'll probably want to get:
(1) Project box (optional)
(1) Antenna (a loop antenna or elevated antenna is recommended and can be made with copper wire)

The tools you'll need are:
Soldering Iron/ Solder (optional)
Multimeter
Oscilloscope (http://www.seeedstudio.com/depot/DSO-Nano-v3-p-1358.html?cPath=63_65) (optional)

That's it. Yup, that's all. Once we've got it all, let's begin.

Step 2: Build the Circuit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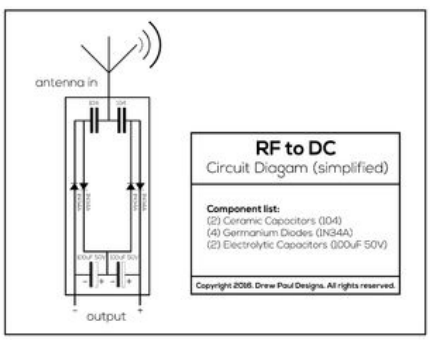
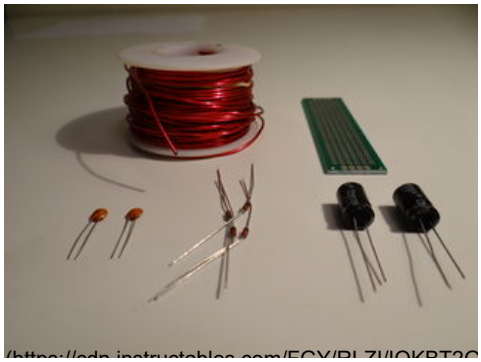
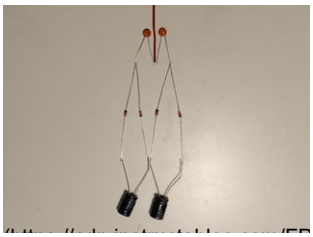
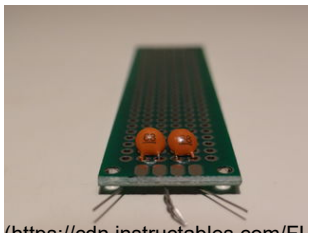
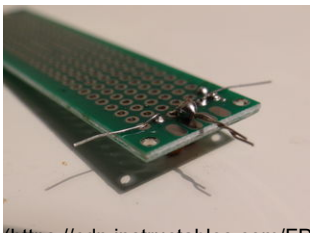
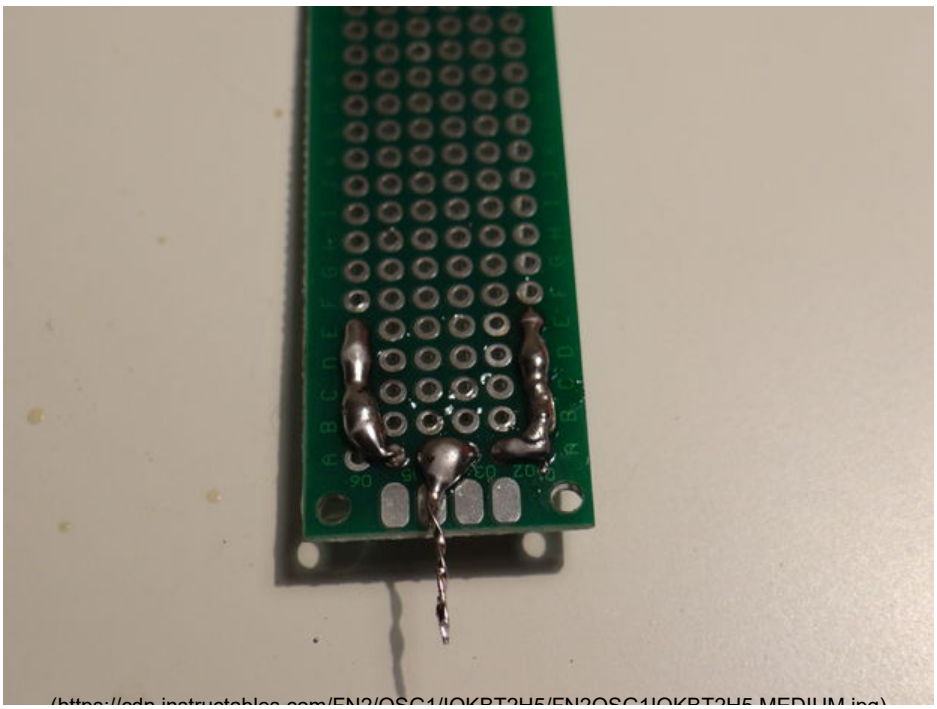
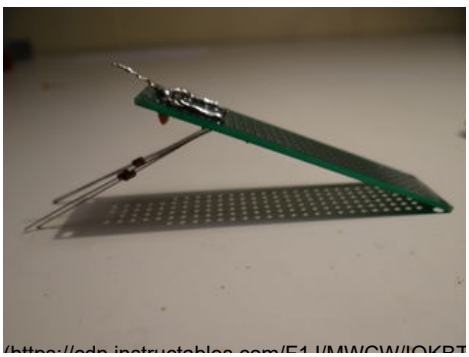
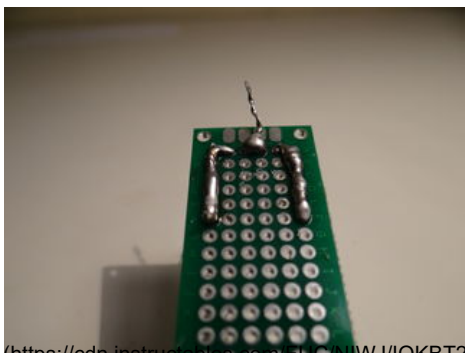
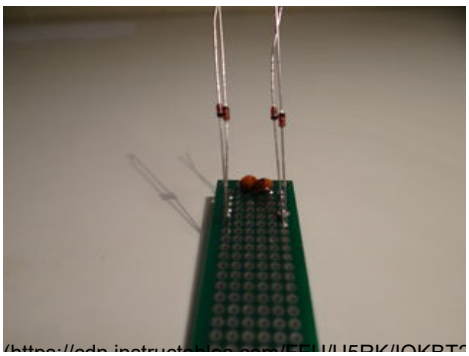
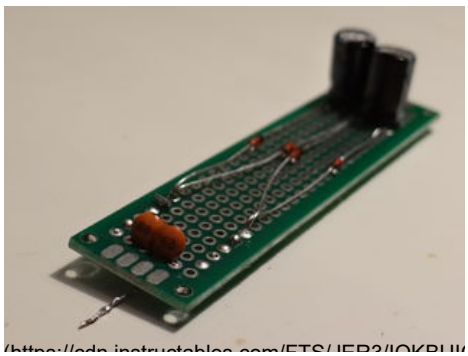




We're going to build the simplest version of this circuit variation in order to understand how each component interacts and as a proof of concept.

There are three simple systems at work in the circuit that are composed of capacitors, which store energy, and diodes that direct it.

Energy in the band of radio waves, among others, will vibrate a wire antenna on an atomic level, sending a discernible signal to its lead. This signal will then

meet the junction between two ceramic capacitors wired in series. This junction will force positive charge from the wave to travel in one direction and negative charge in the other direction which, when collected again, makes the signal uniform and polar. Connecting the two capacitors in series creates leads on each end; the now positively charged side of one and the now negatively charged side of the other creates a two cell battery.

This next stage of the circuit takes a signal with a net value of zero, adds the absolute values of the positive and negative amplitudes with respect to the origin and produces a positive integer. This concept can be thought of as taking:

(+1) + (-1) + (+1) + (-1) = **0**

and converting it to:

| [(+1) + (+1)] | = **2**
+
| [(-1) + (-1)] | = **2**
= 4

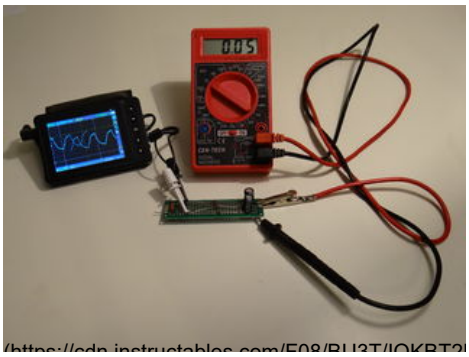
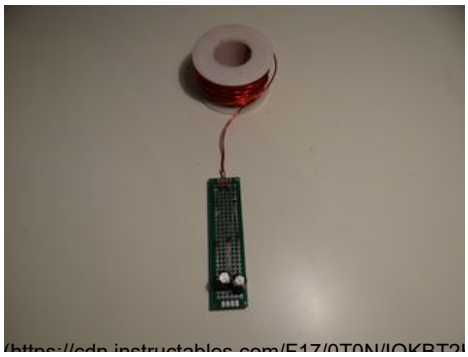
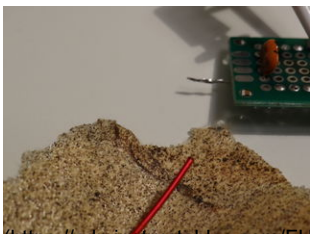
Isn't math fun?

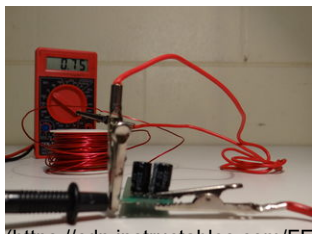
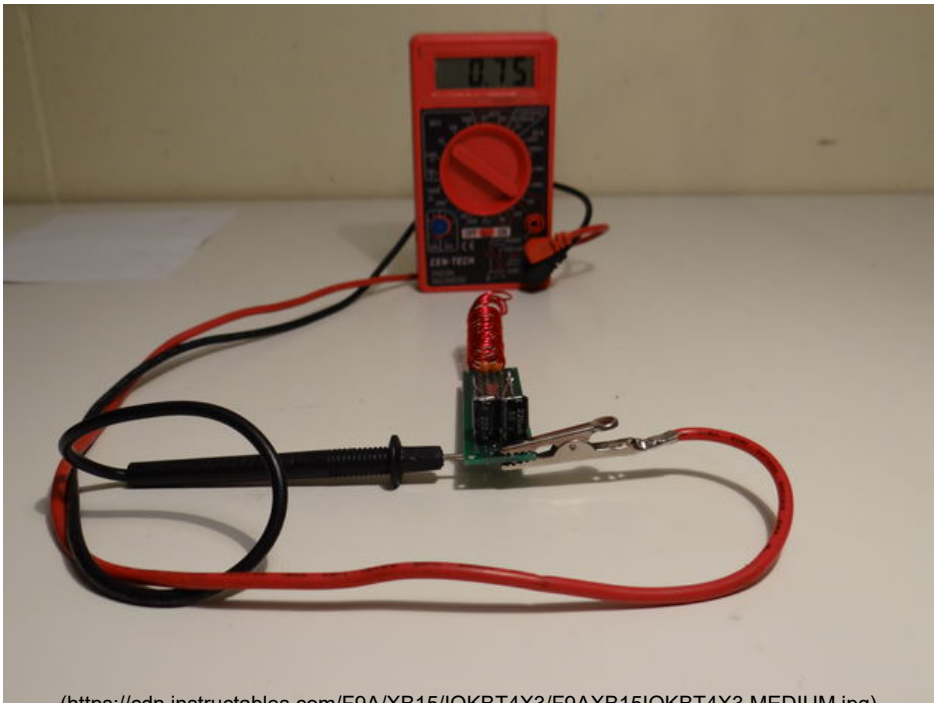
To each of these leads from our two capacitors in series, we will connect two crystal diodes, one facing each direction, to form what is called a bridge rectifier. A bridge rectifier is a configuration which will convert an alternating current to a direct one by cleverly rerouting the signal.

By connecting the bridge rectifier as shown in the circuit diagram, this direct current from the diodes then charges the electrolytic capacitors. This stage normalizes the amplitude, making the current constant and usable.

Components can easily be twisted together for testing and then soldered to a circuit board to secure.

Step 3: Test and Optimize Your Circuit

To test and analyze our circuit, we'll be using a digital voltmeter and oscilloscope.

By connecting a voltmeter to the output, we'll immediately begin to see a small voltage climbing in the 10-100mV range. If not, we'll want to check our connections and make sure the circuit is not isolated from the environment by taking it outside to a clear area.

Then, by connecting an oscilloscope to the outside leads of our two ceramic capacitor bank, we will see the the polarized signal being captured from the air around us. We can then connect after the diodes to see our varying direct current and then to after the electrolytic capacitors to see a normalized, usable direct current at our output.

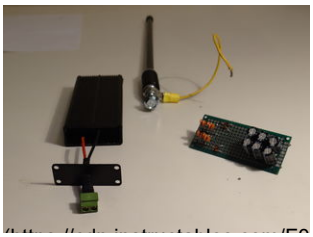
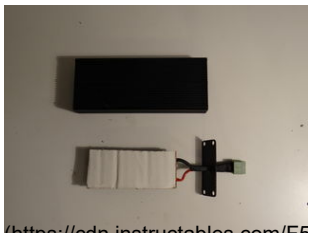
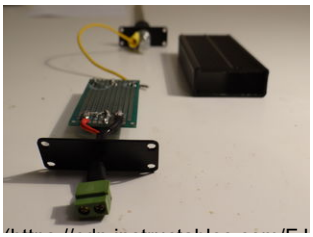
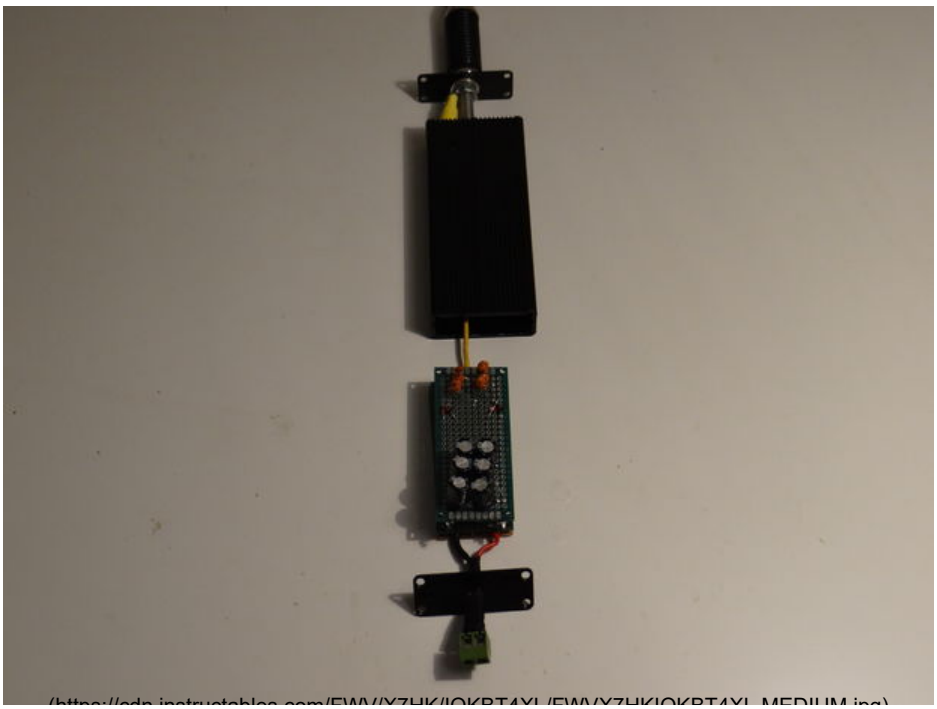
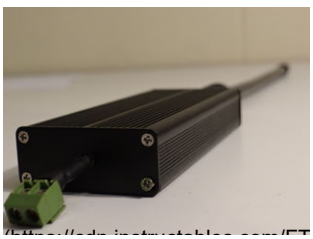
We can then optimize the input resistance in two ways. Firstly, we can add additional ceramic capacitors in parallel to our original two and make sure our soldered connections are consistent and thick in this area.

We can optimize the circuit's capacity by adding electrolytic capacitors in parallel to our original two which will allow this circuit to charge slightly when not in use. For this purpose, a charging circuit can also be added here in order to incorporate an optional battery bank.

We can optimize the antenna by attaching loops and coils of copper wire in various positions, store-bought antennas or by stringing some wire up to the highest point you can reach.

We don't have to stop there, either. We can now connect multiple circuits in series to increase voltage or in parallel to increase current. This can be done indefinitely.

Step 4: Add a Case and Antenna

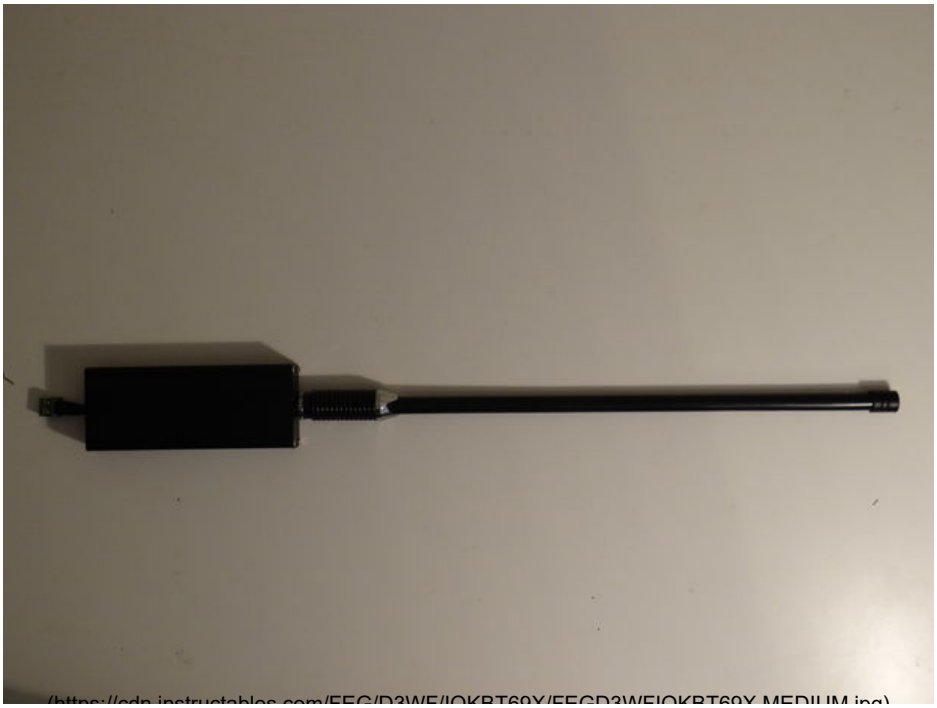






After choosing an antenna in the last step we'll now want to permanently wire it. Whether you choose a compact antenna for portability or a tall fixed antenna for power and range, we will wire it in the same manner according to the diagram in the previous step. Note that the input on the configuration here is grounded to the metallic case, and thus the users hand, and incorporation of longer antennas will require proportionally more substantial grounding.

We will then attach a terminal to the output to allow us to connect this circuit to an electrical device or charging circuit and battery bank.

Next, we will add a case, making sure to isolate exposed leads with non-conductive material especially if mounting in a conductive case. A piece of cardboard secured with glue is sufficient for the circuit's bottom and shrink wrap or electrical tape can be used in the case of any additional exposed leads. Drill two holes in your enclosure, one for the antenna or antenna lead and another for your output terminals. You can then insert your components, fasten the enclosure and your device is ready to use!

Step 5: Your Crystal Energy Receiver Is Complete!



Your Crystal Energy Receiver is now complete and ready to use!

I built a portable version, for proof of concept and demonstration purposes. However, you can go as big as you want- to passively charge batteries or run equipment remotely; or go as small as you want- to power sensors, RFID devices, small electronics and more.

I used this harvested energy to easily power a low-consumption quartz clock, a digital chronograph with integrated circuits and LCD and was even able to momentarily rotate a small dc motor.

Because of its simplicity this device is a durable, efficient and reasonably effective method of harvesting radiant energy in a simple, replicable and sustainable way. I humbly hope that the contributions made here, and by those

reading, can be one day used by people worldwide to conveniently capture free energy.

Thanks for checking out my project and I look forward to seeing everyone's variations, suggestions and improvements!

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Comments

be nice We have a be nice comment policy. Please be positive and constructive.

I Made it! Add Images Post Comment

TzefaiO (/member/TzefaiO/) 2017-09-30 Reply

Can you ship to Hong Kong when I bought it? How can I contact you?

VinyasiQ (/member/VinyasiQ/) 2017-08-30 Reply

http://is.gd/battchar2

http://is.gd/portablebatterycharger

Dornier (/member/Dornier/) 2017-07-29 Reply

1: what rating capacitors (ceramic)?

2: what rating capacitor (electrolytic)?

3. why both types of capacitors and not one type?

4. You've mentioned grounding to earth, but where is the grounding line?

5. Looking to make this a kids' project to charge a phone so expanding these size isn't a problem at all.

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ Dornier (/member/Dornier/)

2017-07-29 Reply

Thanks for checking out my project. A lot of these questions are covered in the Instructable but I will answer them briefly for you.
The ceramic capacitors used are 104s and their purpose is to polarize the charge. Electrolytic caps are 50v rated and are used to hold the charge in order to provide a usable current. In this example the ground is connected to the case, and thus grounded through the users hand. Scaling this up for that purpose will require a more significant antennae, ground and wiring up multiple circuits in series which is covered in more detail in the article. Let me know if you have any other questions and be sure to post your project pictures here when you're done for everyone to see!

Loveforcats (/member/Loveforcats/) 2017-07-17 Reply

Hey,how much maximum voltage can be obtained at the output by adiing capacitors in parallel....????

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ Loveforcats (/member/Loveforcats/) 2017-07-17 Reply

Thanks for checking out the project. Great question. Adding capacitors will only increase the capacity (mAh). If you want to increase voltage you can build multiple circuits in parallel and make sure your antennae is as tall as possible.

Loveforcats (/member/Loveforcats/) ▸ DrewPaulDesigns (/member/DrewPaulDesigns/) 2017-07-26 Reply

I dont want to place my antenna on roof top.I want a min of 5 volt and max of 20 volts and a portable circuit.So what shud i add to the circuit for getting it.??? and what kind of antenna shud i make..??

Loveforcats (/member/Loveforcats/) ▸ Loveforcats (/member/Loveforcats/) Reply 2017-07-26

Can u please send me a video of how your project is working..??

Loveforcats (/member/Loveforcats/) ▸ Loveforcats (/member/Loveforcats/) Reply 2017-07-17

adding*

galihprog (/member/galihprog/) 2017-07-14 Reply

sir, why the first kit show value 0.75v and the second kit show value 3.50v,

what is different make? (coil turn, coil component, total component, etc..)

Raghdado (/member/Raghdado/) 2017-07-04 Reply

What a nice instructable! But I have few questions though, if you don't mind.
How long did it take you to complete the project?
How much did it cost you?
Would it be realizable for the duration of one month only (for a beginner)?
And most of all, Could you tell me about the output (after testing)?

Thank you :)

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ Raghdado (/member/Raghdado/)

2017-07-04 Reply

Thank for checking out my instructable! This project can be done in an afternoon and the kit can be found here: http://www.drewpauldesigns.com/crystal-energy-receiver-kit.html

Before getting started you will also want to get a soldering iron and take some time to practice with it.

Just follow the instructable carefully and you will have no problem.

You can get everything you need here: http://www.drewpauldesigns.com/crystal-energy-receiver-kit.html

Raghdado (/member/Raghdado/) ▸ DrewPaulDesigns (/member/DrewPaulDesigns/)

2017-07-04 Reply

Thank you, I can't wait to start it :)

But before, I just want to know the testing result (voltage, current...)

alaafprojs (/member/alaafprojs/) 2017-01-22 Reply

Cool :-)

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ alaafprojs (/member/alaafprojs/)

2017-01-22 Reply

So cool, right? Give it a try! Its easy, fun and the results are astounding!

alaafprojs (/member/alaafprojs/) ▸ DrewPaulDesigns (/member/DrewPaulDesigns/)

2017-05-30 Reply

The problem is I don't have any germanium diodes

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ alaafprojs (/member/alaafprojs/)

2017-05-30 Reply

You can get the kit here:

http://www.drewpauldesigns.com/crystal-energy-receiver-kit.html

alaafprojs (/member/alaafprojs/) ▸ DrewPaulDesigns (/member/DrewPaulDesigns/)

2017-06-01 Reply

Yes, I know.

But I just got some germanium diodes and I have all of the other parts

alaafprojs (/member/alaafprojs/) ▸ alaafprojs (/member/alaafprojs/)

Reply 2017-06-01

I just made it and I tried it with germanium diodes and 1N4001 diodes and the 1N4001 diodes worked better

PS: I have found some other circuits that work better

Raghdado (/member/Raghdado/) ▸ alaafprojs (/member/alaafprojs/) Reply

2017-07-03

Could you please post the image or the link of the other circuits that work better ?

---

frisbrob (/member/frisbrob/) 2017-03-30 Reply

I am always looking for ways to generate power, mostly because it's fun but most of our renewable energies have their down sides. Solar only works while it's day time, wind only works while there is wind. I don't know about where you live but here in Sandy Utah there is no sun at night and at night during winter or summer there is no wind at night, so that takes care of generating power at night. I have looked at several of these pulling power out of the air things, I have even looked at finding Tesla's plans and there is nothing usable. I don't know how to do the math to figure out what parts are needed to build these diy collectors and generators. To me it sounds like you would need an entire back yard to hold anything big enough to do any good, most of the things people build might charge a cell phone but that's about it. If I was to build this unit but big enough to constantly run an 800 watt unit, how big would the antenna need to be and what parts would I need? Just curious.

---

medicinet1 (/member/medicinet1/) 2017-03-26 Reply

Free Energy Generators

medicine turtle cherokee it a great ideal

http://www.hal-lo.at/

PhillipH50 (/member/PhillipH50/) 2017-03-12 Reply

I think what you should look into is tuning this simular curcuit to recieve in the frequency range of 7.83hz to 8.4hz. Because this is the earths magnetical resonance frequency. Tesla's paper state that 1 meter = 100v on a horizontal wire. So at 3 feet tuned into this frequency can produce 100v and at 6 feet produce 200v, ect..
About 10 years ago there was a video that was taken down of a Russian man who was demostrating his setup with a hand full of scientists. I don't remember all the details but one thing I do remember was that his antenna was 68 feet of 14 gauge copper wire hanging 6 feet above the ground horizontally. The ground stake was 1/2" rebar 6' into the ground.

You can look into these frequencies by searching "Schumann resonances are global electromagnetic resonances in the cavity formed between the earth’s surface and the ionosphere. "

Just some of my thoughts..

---

tatebullrider (/member/tatebullrider/) 2017-02-09 Reply

Great instructable!! Just a quick question though. So the higher the antenna is in the atmosphere the better? And if so, you could (in theory) use a long thin wire attached to a kite or a weather balloon to temporarily generate more power, correct?

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ tatebullrider (/member/tatebullrider/) 2017-02-09 Reply

You got it. You could use a kite or even a drone or quadcopter to take measurements at different altitudes. Post your data for us to see!

16me121lucifer (/member/16me121lucifer/) 2017-01-07 Reply

Sir i am getting 0.06v but negligible current. I am unable to light a LED. I wanted to know how can we connect similar circuits in series or parallel. I would be grateful if you upload a schematic diagram for the same.
Thank you.

16me121lucifer (/member/16me121lucifer/) 2017-01-02 Reply

Dear sir,
Can i use antenna of my old radio set to complete the given project?

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ 16me121lucifer (/member/16me121lucifer/) 2017-01-02 Reply

Yes, you may. In fact, that is what it was designed for; you'll just using that energy for a different purpose. From there, the taller the antennae you use, the better the results.

Alijo0nn (/member/Alijo0nn/) 2016-12-30 Reply

Dear Paul

Is it possible that this circuit generates 1 ampera current with highest place of Antenna and suitable Earth?

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ Alijo0nn (/member/Alijo0nn/) 2016-12-30 Reply

A single circuit, no. However, by running these circuits in parallel current can be increased to suit your output needs.
Thanks for checking out my Instructable and if you have any other questions feel free to let me know.

16me121lucifer (/member/16me121lucifer/) 2016-12-21 Reply

Sir can we light led with help of this project? & Can we do something so that the received energy is used only when we require? ie. something like switch?

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ 16me121lucifer (/member/16me121lucifer/) 2016-12-21 Reply

Yes and yes. Give it a try and let us know how it works out.

Alijo0nn (/member/Alijo0nn/) 2016-12-16 Reply

Dear Pual

How can i connect some these circuits together?

is it work, if i connect 10 matched circuits to an antenna?

Best Regards

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ Alijo0nn (/member/Alijo0nn/) 2016-12-16 Reply

You sure can. Give it a try and post your pictures for us to see.

cmlucht (/member/cmlucht/) 2016-11-23 Reply

what if you hooked up some joule thief's low input higher output?

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ cmlucht (/member/cmlucht/) 2016-11-24 Reply

Yes, try it.

vvince (/member/vvince/) 2016-11-13 Reply

p.s

good ible all we need now is a nano amp amplifier

vvince (/member/vvince/) 2016-11-13 Reply

seems to me that thomask19 works for some big energy company who have suppressed ALL FREE ENERGy

egads man we cant have the minions getting ANYTHING free

there's no PROFIT in that

SHOE0007 (/member/SHOE0007/) 2016-11-07 Reply

I not trying to be rude but how do you get free energy from nothing? Unless you do this under a high voltage power line (very dangerous) i am not sure if this would work?

In state matter is neither created or destroyed so were is this energy coming from??

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ SHOE0007 (/member/SHOE0007/) 2016-11-07 Reply

Not at all and this is a very good question.

As Faraday and others have demonstrated, magnetic bodies in motion produce an electric charge. If you've ever used a magnetic compass, you know that the earth, along with orbiting bodies, are magnetic with a north and south pole. These moving bodies induce a charge in our atmosphere. This charge, with our new Crystal Energy Receiver, can be captured and harvested as usable energy.

Thanks for reading. Give it a try and be sure to follow and visit www.DrewPaulDesigns.com for more projects.

Brandonoutdoors (/member/Brandonoutdoors/) 2016-11-04 Reply

i like this "ible". I'm seeing a lot of negative comments though so I want lend my support to the author.
In no way is converting radio waves into electricity stealing. I mean think about it. When you turn on a radio to listen to music nobody bills you for the radio waves. They are paid for by ads.
With a homesteader in mind who may not live anywhere near transmission lines could use somthing like this at a larger scale (our planet is saturated in rf). when radio waves cross your property line they become yours to do with what you will (in my opinion). I say build (or buy) a large antenna and start experimenting with a larger scale.

cloethix (/member/cloethix/) 2016-08-02 Reply

good project

rah187 (/member/rah187/) 2016-07-22 Reply

Cool project! I have favorited this instructable and it is possible that I will eventually construct it. This circuit is actually very simaller to a few others I have seen on the internet, all of which produce around three volts. I have even seen people use crystal radio sets to generate electricity for LEDs. Of course, there are limitations to this setup, including the very low current and the fact that making this unit much larger than it already is may be considered illegal. The truth is, nobody would probably care about one device running off of the RF, seeing as seemingly innoccuous grounded objects such as flagpoles can also "steal" radio waves. However, I am glad I read through the comments before making a gigantic fractal antennae and getting myself in trouble.

İlhamiA1 (/member/%C4%B0lhamiA1/) 2016-07-18 Reply

Lolz its ??? Nano amperes and 0.06 v

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ İlhamiA1 (/member/%C4%B0lhamiA1/) 2016-07-18 Reply

Eureka! You got it working! You now have proof of concept. Now it's time to increase your current.

First, try extending the antennae all the way to the highest point you can find (try your roof). Then connect a ground lead to the junction between your output terminals (between the two electrolytic caps in series) and run that to a piece of steel rod pressed into the dirt.

Once that is complete, you can optimize it even further! Now, with your antennae and ground set up start wiring in more circuits. Wire them in series, parallel or both according to your needs. Put all of the components on the same circuit board for a supersized version.

Thanks for checking out my 'ible! I hope this helps!

unbottled19208306 (/member/unbottled19208306/) 2016-07-10 Reply

When you say the 2 12 and 6 matched caps, do you mean 12 and 6 volt caps?

DrewPaulDesigns (/member/DrewPaulDesigns/) ▸ unbottled19208306 (/member/unbottled19208306/) 2016-07-11 Reply

The voltage capacity is variable, however it is important to use matching pairs of both ceramic and electrolytic capacitors. Thanks for checking out my project and let us know how yours comes out!

Arghus (/member/Arghus/) 2016-06-23 Reply

can u get it to 5v? for a usb port?

RowanS1 (/member/RowanS1/) ▸ Arghus (/member/Arghus/) 2016-07-01 Reply

yes its possible but current would be too low to be usable

RobertH7 (/member/RobertH7/) 2016-06-13 Reply

Congratulations! You built a full-wave rectifier! You passed the first chapter of basic electronics.

Unfortunately, this particular rectifier is an extraordinary waste of time. It produces a negligible amount of power. Your oscilloscope voltage reading is meaningless. Power is voltage TIMES current, and something like this will produce current in the range of nano-Amperes.

Just get some solar cells and call it a day. They absorb energy from the peak of the sun's blackbody radiation spectrum (which happens to be visible light), not in the negligible radio frequency range. The cheapest, least efficient solar cells out there will produce orders of magnitude more power.

More Comments

Free Energy From Thin Air! by DrewPaulDesigns (/member/DrewPaulDesigns/) in science (/technology/science/)

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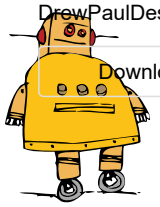

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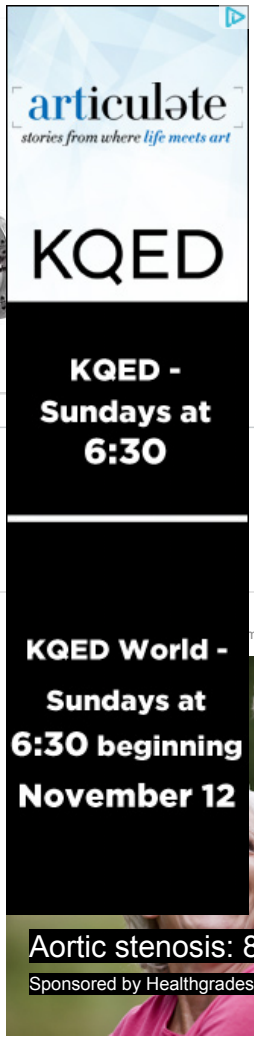

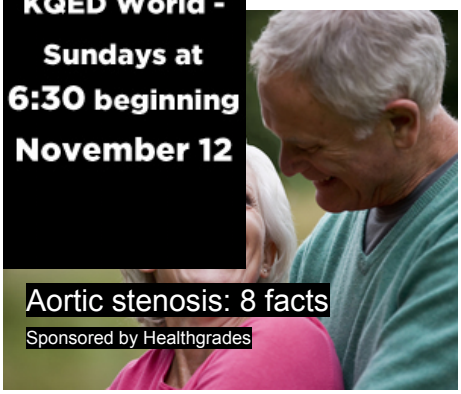

Probe for measuring in the Radio Frequency range built inside an earphone jack, designed to have minimum capacitance and complete shielding.

Step 1: The Parts

About This Instructable

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neelandan (/member/neelandan/)
it2n.net (http://it2n.net)
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Bio: I'm Chandra Sekhar, and I live in India. I am interested in electronics, and building small one-off circuits around tiny chips (the electronic kind).

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You will need:

A diode. Germanium point contact, OA79 or 1N34 is the preferred and traditional choice. But a modern substitute will be a schottky (gold bonded) diode. These have a low forward voltage, below 250 millivolts or so, versus the 600 or so of the silicon diode.

A few capacitors, chip type, value not critical, around 1 nf to 100 nf. (or 1000 pf to 0.1 microfarad)

One resistor, 1 Megohm.

One stereo earphone jack.

One pin from a turned pin IC socket. This forms the 'hot' end of the probe, so that various pins can be inserted for probing. A short piece from a needle for exploratory work. A piece of wire soldered to the circuit board when using it for a tuning up session, requiring it to be in place for a long time.

The circuit is traditional, as below:

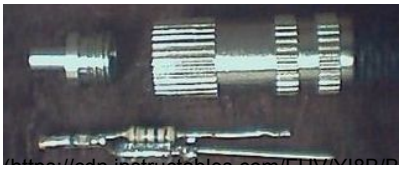
```
   C1          R<--||-------/\/\/\/\-------------o +      |               |
---              |      ^               |     /-\ diode        === C2      |
             |      |               |<--------------------------------o -
```

Step 2: Connect It Up

I used a small sliver of circuit board to hold C1. The turned pin socket was fitted to one end, the capacitor C1 soldered to it and then the resistor and diode leads formed to shape and soldered to the capacitor. It is shown next to the stereo earphone jack in order to see whether it will fit. Adjust sizes, get a smaller capacitor etc until it does.

The socket, capacitor and supporting board will go inside the barrel of the jack.

Step 3: Take the Jack Apart



A bit of filing, twisting and pulling will result in the jack coming to pieces. It is composed of a large number of precision formed metal and plastic pieces.

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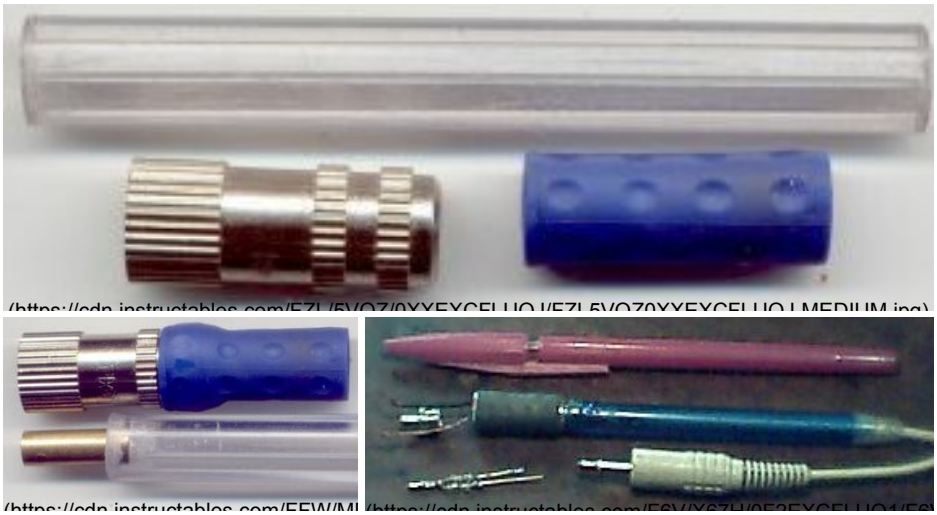
We need just the outer shell and the cover - It is best that the cover of the jack is metal, otherwise the shielding will not be perfect.

Step 4: Form the Body



The body of the probe is made from an old ball point pen. A piece of brass tube was found to serve as a join between the shell of the stereo jack and the ball pen body. The rubber grip from another pen was then slid over the metal to insulate it from accidental contact with circuit parts.

A molded mono earphone lead was used to get the signal out from the probe.

Step 5: Put It All Together

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I used teflon tape to insulate the turned pin socket and capacitor, which had to go inside the barrel of the jack. A dab of petroleum jelly kept its bore from filling up as superglue was applied to hold it in position.

A HAM friend of mine measured about 5pf of capacitance for this probe, which is pretty good.

In use, a sharp probe is inserted into the turned pin socket. The ground lead is wire wrapped around the barrel of the jack. This is necessary if accurate measurements are to be made in a noisy environment.

I hope this will be of use to the HAMs among us.

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RF Probe by neelandan (/member/neelandan/) in wireless (/technology/wireless/)

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I Made it! Add Images Post Comment

Jimmy Proton (/member/Jimmy+Proton/) 2010-12-05 Reply

i never knew what those clear things with the colors on them are until now, why do i only see them on tvs and monitors?

S33PlusPlus (/member/S33PlusPlus/) ▸ Jimmy Proton (/member/Jimmy+Proton/)

2011-07-17 Reply

What kinds of things do you take apart? I find so many signal diodes (1N4148's typically) in trashed electronics I don't bother desoldering them anymore. (Besides, they're cheap enough that using a new one in a project is less troublesome than using a damaged scrap bin diode.)

Jimmy Proton (/member/Jimmy+Proton/) ▸ S33PlusPlus (/member/S33PlusPlus/)

2011-07-17 Reply

I take apart everything that i can get my hands on. =)

S33PlusPlus (/member/S33PlusPlus/) ▸ Jimmy Proton (/member/Jimmy+Proton/)

2011-07-19 Reply

Me too :3
I prefer early 90's or late 80's era devices, before everything went surface mount.
If you take recent products apart, you'll be quite disappointed at the component yield. You've probably seen few discrete signal diodes because manufacturers cram everything they can into ICs to reduce component counts.
On the other hand, an early 90's handheld four-in-a-row style game I recently took apart (lack of backwards current protection fried a LED) has 34 discrete transistors in it!
Try going through free piles after garage sales or the free section on your local craigslist boards.

Jimmy Proton (/member/Jimmy+Proton/) ▸ S33PlusPlus (/member/S33PlusPlus/)

2011-07-20 Reply

Old CRT TV's probably have the best stuff in them.
Whenever I take apart new things I put the boards in the toaster oven or i destroy them.

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taterkiller (/member/taterkiller/) 2011-01-06 Reply

will this work as an antenna

PetervG (/member/PetervG/) 2007-02-09 Reply

I thought the first picture was a person trying to pull their bike away from a huge cylindrical crusher..

dcshoeco33 (/member/dcshoeco33/) ▸ PetervG (/member/PetervG/)

Reply 2007-02-09

lol i seriously rofl'd it does look like that

chrisayad (/member/chrisayad/) ▸ dcshoeco33 (/member/dcshoeco33/)

Reply 2008-06-30

lolool

PetervG (/member/PetervG/) ▸ chrisayad (/member/chrisayad/) 2008-06-30 Reply

Wait, how does it look like that. I wish I knew what I meant.

Punkguyta (/member/Punkguyta/) 2007-02-18 Reply

Oh wow, this does look good. Is it as fragile as it looks?

neelandan (/member/neelandan/) ▸ Punkguyta (/member/Punkguyta/) Reply

2007-02-19

It isn't, because the socket at the tip, the series capacitor and the ends of the resistor and diode are held securely inside the tip of the audio jack. You may use hot melt glue to hold them in position. I wrapped it in teflon tape (available at plumbing stores) and secured it in position using superglue. Epoxy would also work.

Punkguyta (/member/Punkguyta/) ▸ neelandan (/member/neelandan/) Reply

2007-02-19

I think that epoxy would be best, hot glue expands, and you tend to get a but too much, super glue is runny and I always get it on my fingers (my tongue too once, don't ask).

cylver (/member/cylver/) 2007-02-11 Reply

Okay, for those of you who want a more in-depth understanding of how it works... c1 is a dc-blocking capacitor. It's there to make sure you're only measuring the AC voltage in the circuit. d1 is there to convert the ac into dc pulses. c2 is a filtering capacitor, so that you're reading the peak voltage (or close to it) rather than the average, which will be quite a bit lower. It also allows any remaining ac a nice path to ground. R makes sure that you don't draw too much current through your meter, destroying it. One thing to keep in mind is that you'll probably want a shielded cable, to keep from picking up stray AC from the board you're testing. RG 174 coax would be nice if you've got some. Connect the shield to the - side of the circuit, and the center conductor to the + side. Slap a BNC connector on the end, rather than a phone plug, and you can use the probe with an oscilloscope, too.

DonTron (/member/DonTron/) 2007-02-10 Reply

This is a classic from the AARL. The neatest ones were built, IMHO, out of lipstick tubes. The non-linearity of the diode, not to mention the integrator, limits the accuracy of this technique, so nowadays, something built out of today's high-speed op-amps might be possible. A bolometer would be more complex, and slower, but far more accurate. This is a very useful instrument, where it is indicated. "Rectify and integrate" is also the principal behind simple AM detection, although with a shorter time-constant, typically.

admanrocks (/member/admanrocks/) 2007-02-09 Reply

this may be a stupid question. but i dont know what it does. so what does it do?

sysadmn (/member/sysadmn/) ▸ admanrocks (/member/admanrocks/) Reply

2007-02-09

Most voltmeters don't accurately measure AC (alternating current) at high frequencies (RF = Radio Frequency). At RF frequencies, the components in the probe convert the AC value to an equivalent DC

value that a voltmeter can accurately measure. There is a good writeup here (http://www.io.com/~n5fc/rfprobe1.htm (http://www.io.com/~n5fc/rfprobe1.htm))

admanrocks (/member/admanrocks/) ▸ sysadmn (/member/sysadmn/) Reply

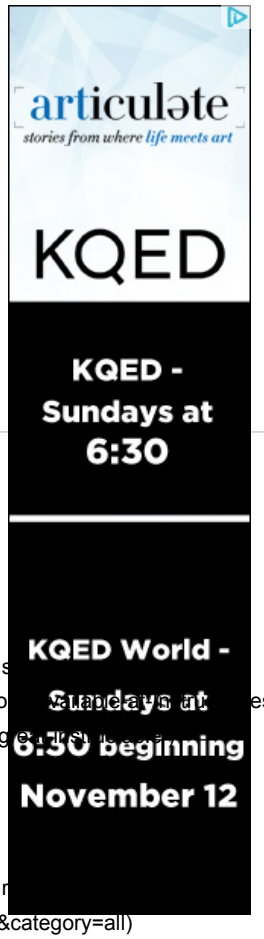
2007-02-10

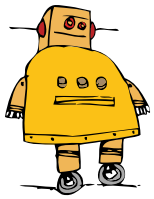
ok, cool

robonut625 (/member/robonut625/) 2007-02-09 Reply

Nice instructable, neelandan.... one of those pieces of test equipment you don't think you need until you have one!

More Comments





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END FED 6–40 Meter Multiband HF Antenna

Introduction

This EARC project produces an inexpensive, multiband, end fed HF antenna matchbox that is quick and easy to setup and use. The end fed feature adds portable convenience, but does present another issue. The problem with an end fed half wave antenna is that the antenna presents high impedance, creating a significant miss match with the usual transceiver impedance of 50 ohms. This miss match is significantly greater than typical tuners can accommodate without a matching transformer.

This project creates a trifilar wound, 9:1 UNUN (unbalance to unbalance) toroid matching transformer that will match the high input impedance of an end fed antenna into the range where most antenna tuners can produce good performance. The matchbox handles 100 watts of power. ***This project requires an antenna tuner to achieve satisfactory SWR***. If you need an auto tuner, we recommend the LDG line of auto tuners, shown to work well with these matchbox antennas, and highly rated by users. They are reasonably priced and available at most ham stores.

The matchbox project uses readily available common hardware and materials listed below.

Matchbox Parts List

1	small plastic enclosure (shown right)
1	powdered iron toroid T130-2
3	20” pieces of 18 AWG solid insulated copper wire in red, green, and black
2	# 8-32 x 3/4” hex head machine screws
2	# 8 lock washers
2	# 8 ring wire lugs
2	# 8 flat washers
2	# 8 lock washer/nut combination
2	# 8 wing nut
1	SO-239 panel mount connector to fit keyed enclosure opening

30’ # 18 AWG insulated stranded wire antenna with ring lug attached Small amount of clear silicone caulk to secure toroid in place

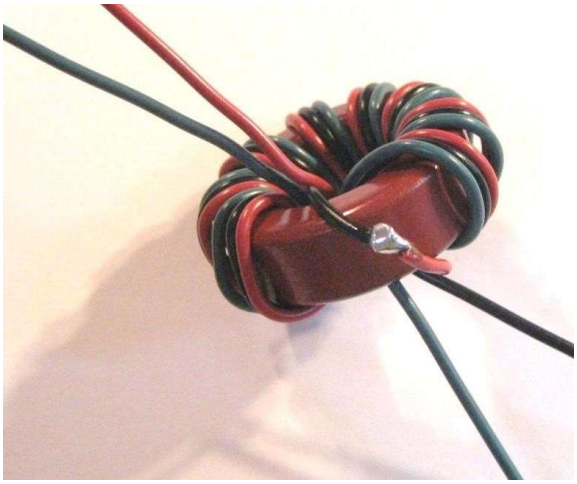
Preparing the MATCHBOX Plastic Enclosure

The enclosure needs one 5/8” hole for the SO-239 connector, and one 11/64” hole for the counterpoise wing nut on the lower side of the enclosure. An 11/64” hole on the upper right side of the box facilitates antenna connector mounting.

TOROID WINDING

First, wind the three 20” pieces of insulated solid wire onto the toroid. Place the wires as shown green-black-red, and wrap the toroid 9 turns so that it looks like the photo on the right. Notice there are three wires extending from the left winding and three wires extending from the right winding. As the connections are completed, the steps refer to the specific wires by left or right and color.

NOTE: Count turns by counting the number of times the wire goes thru the toroid center.



Crimp together and solder ***the left black wire*** with the ***right red wire*** When the step is completed, it will look like the right photo.

The next three steps should appear as shown in the first photo on page 3. Crimp and solder a # 8 lug to the ***left red*** wire about 2” from the toroid. The completed lug will later connect to the antenna connection bolt on the upper inside of the enclosure.

Twist the **left green** wire with the ***right black*** wire. Strip the ends of the two wires and twist together at about 2”. This twisted pair will solder to the center connection of the SO-239 connector in a later step.

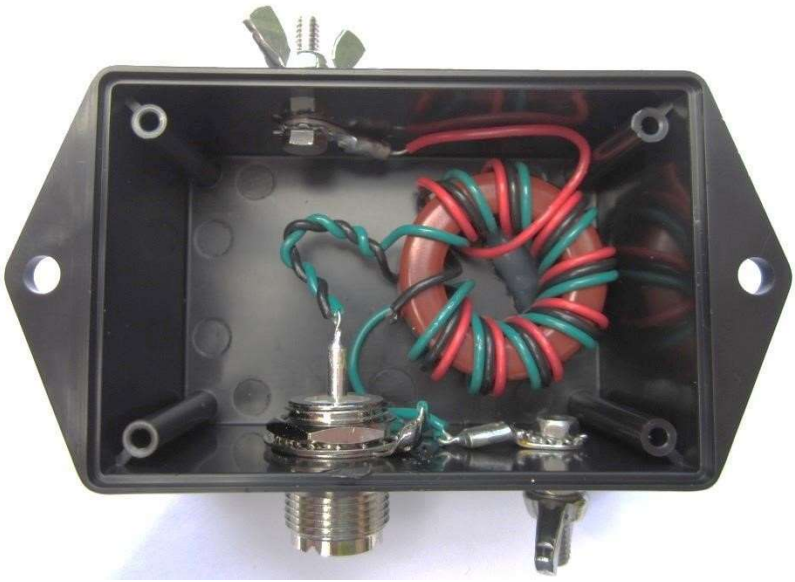
Trim and strip the remaining ***right green*** wire at about 2”. Cut an additional 2” piece of green wire, and crimp and solder both wires to a # 8 lug. The 2” green wire will connect to the ground

connector on the SO-239 already installed in the enclosure. Strip remaining green wire end 3/8” and bend into a hook for connection to the SO-239 ground connector.

The SO-239 connector and ground lug should be installed through the keyed hole in the lower right with the mounting nut securely tightened.

Solder the green and black twisted pair to the SO-239 center connector. Solder the green wire hook to the ground lug on the SO-239 connector.

From inside the box, place an 8-32 machine screw through a lock washer, the # 8 lug on the green wire, then through the lower 11/64’” mounting hole. Place a flat washer on the outside protruding machine screw followed by # 8 lock washer/nut and tighten securely. Place a # 8 wing nut on the machine screw to finish the **counterpoise connector**.



Position the toroid inside the box to allow connection of the red antenna wire lug to an 832 machine screw and lock- washer on the upper box side. Place a flat washer on the outside of the box followed with a # 8 lock washer/nut. Tighten the nut securely. Next, place the wing nut on the **antenna connector** and your project appears as shown above.

A small spot of clear silicone compound is used to secure the toroid from movement in the enclosure. The final assembly step is to secure the box cover in place with four screws.

Preparing the Antenna Wire

Matchbox performance will be determined by two factors: The length of the antenna wire, and the capability of the tuner. The length of the antenna wire should be between 24 and 60 feet for best performance. Additional counterpoise is not normally required, as the coax shield provides the counterpoise function. Wires longer than 60 feet may have excessive impedance for some tuners to properly match. Wires shorter than 24 feet may not radiate as effectively. A 30 foot insulated 18 gauge stranded wire antenna and connecting lug is included with the project and should meet most requirements without need for any counterpoise.

Experience has shown that most external tuners and some internal tuners will tune 80–6 meters with an antenna length of 24’ to 30’. If a longer antenna is desired, the provided antenna can be lengthened.

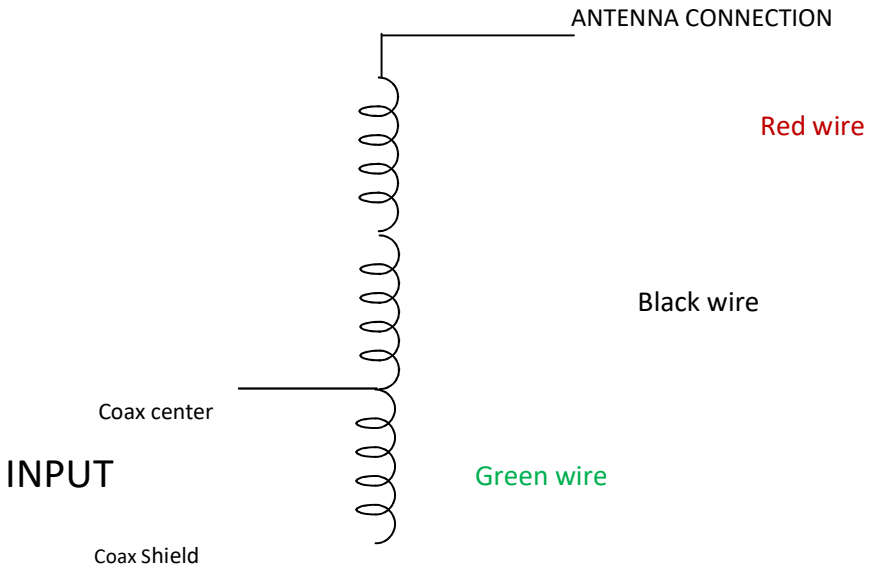
Some tuners, in particular internal tuners, may not tune the full 80-6 meter range. You may need to try different wire lengths to optimize your antenna configuration. If you are having difficulty getting your rig to tune, start with a 26’ wire. This should produce good results on at least 40-6 meters using the narrowest performance range of internal tuners.

Best performance is achieved with a coax of 16’ or longer, with the coax shield providing the counterpoise function. Additional counterpoise is usually not required in this design, although the lower wing nut provides a convenient counterpoise connector if needed. The end fed antenna system works well in horizontal, sloper, and vertical configurations.

Observe established safety practices when working with antennas, and avoid proximity to power or utility wires. Permanent installations should be equipped with appropriate static and lighting protection.

Keep amateur radio safe and fun!

Toroid Wiring Schematic



Feedback on 6-40 End Fed Matchbox Antennas

The antenna is amazing. I receive fantastic signal report from DX stations as well as hams around the country. I previously used a short version GR5V that was in an inverted V configuration on the roof of our house. This antenna was noisy but worked. Your antenna out performs the prior antenna and is much quieter. Thanks again for the fine product that your club offers. ***May 2012***

It took me 10 minutes to install and WOW - within 10 minutes I made my first HF contact (from CT to GA). And in the span of 4 hours made 3 more !! I was transmitting just 5 Watts from Yaesu 857ND using this antenna. This antenna rocks ! Thanks a ton ! 73 ***May 2012***

The 6 - 40 Matchbox antenna works great! Easily tunes with the Z100 tuner - very low SWR's on 20m thru 10m. Antenna was up about 45 feet on tree branch - wire length was about 52.5 feet, slight slope, 55 feet of coax, line isolator at radio. ***Work Hawaii for the very first time. Guess the antenna wanted to call home :-) 57 report, 100 watts, 15 Meter band, Icom 718.*** A great portable antenna that I will use often at Montrose Harbor along the lakefront in Chicago. Thank you for a fine, well made product. ***April 2012***

It arrived today in fine shape. I got is up in the air this afternoon using a 31 foot radiator. Top is at 45 feet. Loads on all bands 80-10M with my LDG auto-tuner absolutely no problem, My first QSO was KH7X in the ARRL SS contest with 100W. Amazing. This thing rocks! It's a great antenna! Nice job and a great buy. ***November 2011***

I set-up my jackite pole today and tried a 68' wire with the 9:1 unun matchbox. I set it up as an inverted L with about 26 feet vertical and 42 feet horizontal. With my IC-703 it tuned 80 through 6 all well below 1.5:1. I am going to leave the antenna up a few days and make a few contacts. I think the matchbox is terrific. ***October 2011***

My 6 – 40 meter end fed arrived through the UK holiday mail and I departed from my norm of a sloper and mounted it vertically on a 10m fishing pole. It is surprisingly effective and rewarded me with a surprise contact on 17m into the Falkland Islands at 20:30UT last night. ***January 2012***
When I attached it to a 10m fiberglass fishing pole and went vertical - wow! Easy 5/9 contacts out to 6,000+ miles when propagation was anything better than the worst. My simple ATU easily matches the antenna with SWR never greater than about 1.4 on any band it is built for. ***December 2012***

You can dither and procrastinate, but with this antenna, you'll be talking to the world in no time on any band that happens to be hot - and with your hard-earned cash hardly touched! ***January 2012***

Just wanted to report back that I have tossed the antenna into a tree about 20' up in an inverted vee shape and I am pleased to report that I can tune anywhere I need on 40-10M using the TS-590 internal tuner!! My first contact was on 40M within 10 minutes of erecting the antenna and was all the way to TX from my condo here in MI using just 10 watts. This is by far the best HAM related investment I have made in a long time. Living in a condo with strict association rules, I am able to conceal this wire in a tree out my bedroom window and talk all over. I'm very happy and would recommend this antenna to anyone in a similar situation. ***March 2013***

I have had your multiband end fed antenna installed here in Singapore for a few weeks now and am very pleased. I used the wire from my previous antenna which is 66 feet long. It works beautifully! I have it hanging vertically outside my 15th story apartment window on a 5 foot horizontal pole (to get it away from the reinforced concrete building). A couple of fish weights on the end keep it from blowing around too much. I would recommend this antenna to anyone who is facing a challenging QTH. ***April 2013*** I received my antenna yesterday, a quick 4 days since I ordered it. I attached it to a 31 foot pole and my IC-706MK2. Within minutes, I made my first contact with 9A4KW in Zagreb, Croatia on 20 meters. Being able to reach out over 4,000 miles gives me real confidence that I have found my portable antenna. Thanks for your rapid response and fine product. ***May 2013***

In only 2 weeks of casual on-air time. If you could see how I had to zig-zag it through a big tree above the roof of my RV, you'd be saying like I am, "how the heck can that work? I worked 90% of the stations I called, and just added Mexico City on 15m that gave me a 5x9. ***September 2013***

Just wanted to let you know how pleased I am with my ERAC matchbox antenna. I put the EARC antenna in the backyard with the feed point at 4 feet and the wire sloping up to 35 feet. I have 125 feet of coax feed line. My transceiver's built-in antenna tuner will adjust the SWR to below 1.2 to 1 on all bands 80-6 meters. Last weekend I worked 65 countries on 5 continents. I live in an antenna restricted community and this antenna has allowed me to enjoy ham radio again. ***February 2014***

I used the EARC Matchbox today. It works GREAT on 20m with a 30 ft radiator in vertical positioning, fed with 25ft of RG175/u coax. I've been talking up and down the east coast with 3W on SSB all afternoon! People ask me to repeat my power output again and again. I can't wait until I get the chance to use it at night, where I think it'll really shine! This has a permanent place in my portable kit. Thanks for the great product! ***April 2014***

I performed a simple install in Virginia making it vertical by using a 28 foot fiber glass kite pole. Within 30 minutes (NO KIDDING) I was receiving Lebanon on 17 meters! I also worked Russia and Eastern Europe that night on 20 meters. I love this antenna and I give it a 100% (5 STAR) rating. ***May 2014***

This little box is magic. NY, IA and NM from GA with ease on 100w ***June 2014***

Just wanted to let you guys know, this matchbox is wonderful. I built the matchbox, connected it to 55 feet of solid #14 about 20 feet in the air. My rig is a Yaesu FT-817ND connected to a MFJ-971 tuner with coax running between the tuner and the matchbox. So far I have worked Slovenia twice, Morocco, St. Helena Island, Cuba, Cost Rica, Vermont, Wyoming, Florida, North Dakota, Brazil, and Colombia. This setup works really well on the higher bands. Thank you for putting this design on the internet. I like it so much its become permanent here. ***September 2014***

This antenna continues to amaze me. It worked well U-shaped "indoors" with 50 watts, and now outdoors as a sloper. It easily loads 10-40m with my TS-570s internal tuner. I'm sold on it and own two of them now. I'm working at least 90% of the DX I chase. I have no RFI in the shack either. ***November 2014***

EARC Antennas are operating in virtually every state, these countries and more!

Argentina Australia Belgium Brazil Canada Chile Denmark Greece

Hong Kong India Ireland Israel Italy Mexico Mozambique New Zealand Russia Singapore Sweden Tasmania United Kingdom

ORDERING A MATCHBOX ANTENNA FROM EARC

Volunteers of the Honolulu Emergency Amateur Radio Club (EARC) carefully assemble the 6-40 End Fed matchbox antennas for those who prefer not to build one. For a donation of **$56** (Including USPS priority mail shipping) club volunteers will build and deliver a 6-40 matchbox with a 30 foot antenna wire anywhere in the U.S., we do not ship internationally. Note that coax cable is not included.

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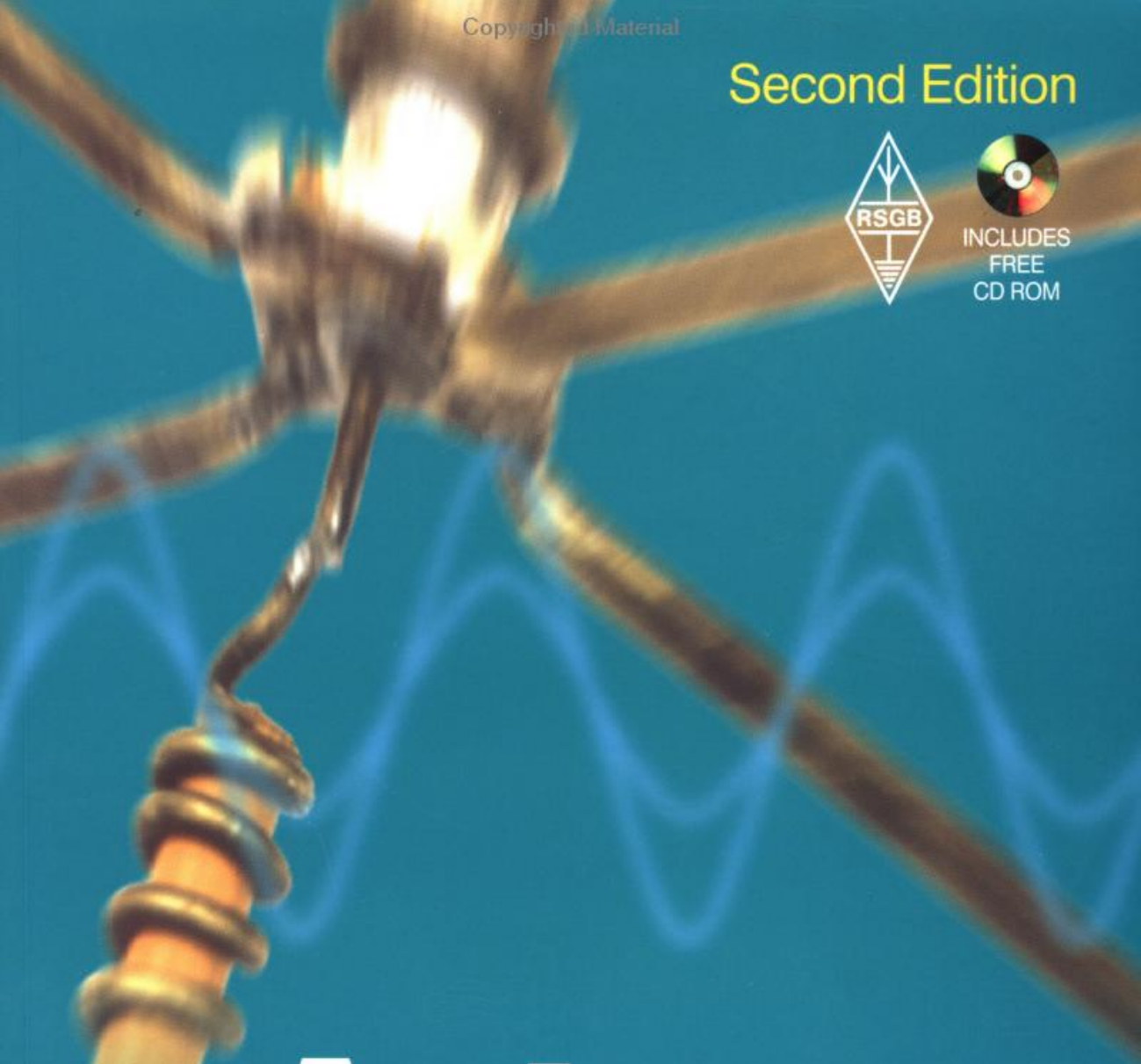

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Preface

If you are interested in amateur radio, short-wave listening, scanner monitoring, or any other radio hobby, then you will probably need to know a few things about radio antennas. This book is intended for the radio enthusiast – whether ham operator, listening hobbyist, or radio science observer – who wants to build and use antennas for their particular requirements and location. All of the antennas in this book can be made from wire, even though it is possible to use other materials if you desire.

These antennas have several advantages. One of the most attractive is that they can provide decent performance on the cheap. As one who has lived through the experience of being broke, I learned early to use bits of scrap wire to get on the air. My first novice antenna back in the late 1950s was a real patched-together job – but it worked really well (or so I thought at the time!).

Another advantage of wire antennas is that they are usually quite easy to install. A couple of elevated supports (tree, roof, mast), a few meters of wire, a few bits of radio hardware, and you are in the business of putting up an antenna. As long as you select a safe location, then you should have little difficulty erecting that antenna.

Finally, most high-frequency (HF) short-wave antennas are really easy to get working properly. One does not need to be a rocket scientist – or professional antenna rigger – to make most of these antennas perform as well as possible with only a little effort. There is quite a bit of detailed technical material to digest if you want to be a professional antenna engineer, but you can have good results if you follow a few simple guidelines.

SOFTWARE SUPPLEMENT TO THIS BOOK

At the time this book was conceived it was noted that the technology now exists to make Microsoft Windows-based antenna software available to readers along with the book. The software can be used to calculate the dimensions of the elements of most of the antennas in this book, as well as a few that are not. There are also some graphics in the software that show you a little bit about antenna hardware, antenna construction, and the like.

ANTENNA SAFETY

Every time I write about antenna construction I talk a little bit about safety. The issue never seems too old or too stale. The reason is that there seem to be plenty of people out there who never get the word. Antenna erection does not have to be dangerous, but if you do it wrong it can be very hazardous. Antennas are deceptive because they are usually quite lightweight, and can easily be lifted. I have no trouble lifting my trap vertical and holding it aloft – on a windless day. But if even a little wind is blowing (and it almost always is), then the 'sail area' of the antenna makes it a lot 'heavier' (or so it seems). Always use a buddy-system when erecting antennas. I have a bad back caused by not following my own advice.

Another issue is electrical safety. Do not ever, ever, ever toss an antenna wire over the power lines. Ever. Period. Also, whatever type of antenna you put up, make sure that it is in a location where it cannot possibly fall over and hit the power line.

The last issue is to be careful when digging to lay down radials. You really do not want to hit water lines, sewer lines, buried electrical service lines, or gas lines. I even know of one property where a long-distance oil pipeline runs beneath the surface. If you do not know where these lines are, try to guess by looking at the locations of the meters on the street, and the service entrance at the house. Hint: most surveyers' plans (those map-like papers you get at settlement) show the location of the buried services. They should also be on maps held by the local government (although you might have to go to two or three offices!. The utility companies can also help.

A NOTE ABOUT UNITS AND PRACTICES

This book was written for an international readership, even though I am American. As a result, some of the material is written in terms of US standard practice. Wherever possible, I have included UK standard wire sizes and metric units. Metric units are not in common usage in the USA, but rather we still use the old English system of feet, yards, and inches. Although many Americans (including myself) wish the USA would convert to SI units, it is not likely in the near future. UK readers with a sense of

history might recognize why this might be true – as you may recall from the George III unpleasantness, Americans do not like foreign rulers, so it is not likely that our measuring rulers will be marked in centimeters rather than inches.* For those who have not yet mastered the intricacies of converting between the two systems:

1 inch = 2.54 centimeters (cm) = 25.4 millimeters (mm)
1 foot = 30.48 cm = 0.3048 meter (m)
1 m = 39.37 inches = 3.28 feet

Joseph J. Carr

*I apologize for the bad play on words, but I could not help it.

CHAPTER 1

Radio signals on the move

Anyone who does any listening to radio receivers at all – whether as a ham operator, a short-wave listener, or scanner enthusiast – notices rather quickly that radio signal propagation varies with time and something mysterious usually called 'conditions.' The rules of radio signal propagation are well known (the general outlines were understood in the late 1920s), and some predictions can be made (at least in general terms). Listen to almost any band, and propagation changes can be seen. Today, one can find propagation predictions in radio magazines, or make them yourself using any of several computer programs offered in radio magazine advertisements. Two very popular programs are any of several versions of IONCAP, and a Microsoft Windows program written by the Voice of America engineering staff called VOACAP.

Some odd things occur on the air. For example, one of my favorite local AM broadcast stations broadcasts on 630 kHz. During the day, I get interference-free reception. But after the Sun goes down, the situation changes radically. Even though the station transmits the same power level, it fades into the background din as stations to the west and south of us start skipping into my area. The desired station still operates at the same power level, but is barely audible even though it is only 20 miles (30 km) away.

Another easily seen example is the 3–30 MHz short-wave bands. Indeed, even those bands behave very differently from one another. The lower-frequency bands are basically ground wave bands during the day, and become long-distance 'sky wave' bands at night (similar to the AM broadcast band (BCB)). Higher short-wave bands act just the opposite: during the

day they are long-distance 'skip' bands, but some time after sunset, become ground wave bands only.

The very high-frequency/ultra high-frequency (VHF/UHF) scanner bands are somewhat more consistent than the lower-frequency bands. But even in those bands sporadic-E skip, meteor scatter, and a number of other phenomena cause propagation anomalies. In the scanner bands there are summer and winter differences in heavily vegetated regions that are attributed to the absorptive properties of the foliage. I believe I experienced that phenomenon using my 2 m ham radio rig in the simplex mode (repeater operation can obscure the effect due to antenna and location height).

THE EARTH'S ATMOSPHERE

Electromagnetic waves do not need an atmosphere in order to propagate, as you will undoubtedly realize from the fact that space vehicles can transmit radio signals back to Earth in a near vacuum. But when a radio wave does propagate in the Earth's atmosphere, it interacts with the atmosphere, and its path of propagation is altered. A number of factors affect the interaction, but it is possible to break the atmosphere into several different regions according to their respective effects on radio signals.

The atmosphere, which consists largely of oxygen (O_2) and nitrogen (N_2) gases, is broken into three major zones: the **troposphere**, **stratosphere**, and **ionosphere** (Figure 1.1). The boundaries between these regions are not very well defined, and change both diurnally (i.e. over the course of a day) and seasonally.

The troposphere occupies the space between the Earth's surface and an altitude of 6–11 km. The temperature of the air in the troposphere varies with altitude, becoming considerably lower at high altitude compared with ground temperature. For example, a $+10^{\circ}$C surface temperature could reduce to -55°C at the upper edges of the troposphere.

The stratosphere begins at the upper boundary of the troposphere (6–11 km), and extends up to the ionosphere ($\approx$50 km). The stratosphere is called an **isothermal region** because the temperature in this region is relatively constant despite altitude changes.

The ionosphere begins at an altitude of about 50 km and extends up to 500 km or so. The ionosphere is a region of very thin atmosphere. Cosmic rays, electromagnetic radiation of various types (including ultraviolet light from the Sun), and atomic particle radiation from space (most of it from the Sun), has sufficient energy to strip electrons away from the gas molecules of the atmosphere. The O_2 and N_2 molecules that lost electrons are called **positive ions**. Because the density of the air is so low at those altitudes, the ions and electrons can travel long distances before neutralizing each other

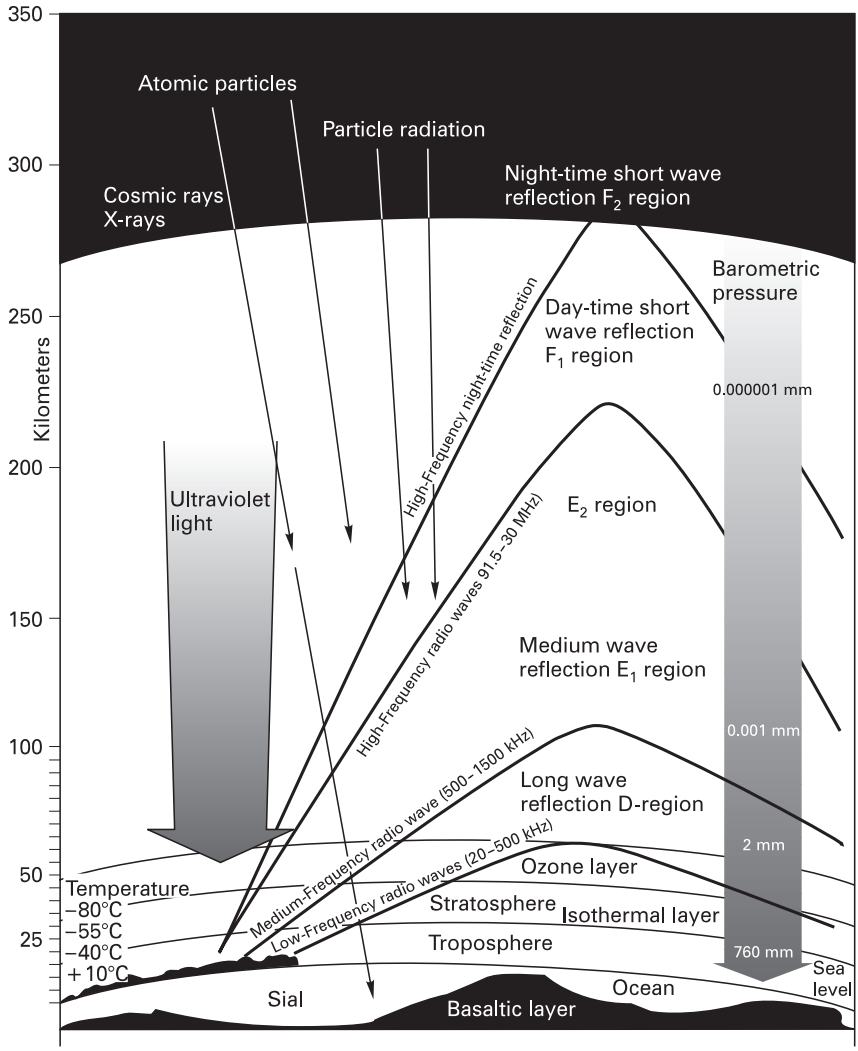


FIGURE 1.1

by recombining. Radio propagation on some bands varies markedly between daytime and night-time because the Sun keeps the level of ionization high during daylight hours, but the ionization begins to fall off rapidly after sunset, altering the radio propagation characteristics after dark. The ionization does not occur at lower altitudes because the air density is such that the positive ions and free electrons are numerous and close together, so recombination occurs rapidly.

PROPAGATION PATHS

There are four major propagation paths: **surface wave**, **space wave**, **tropospheric,** and **ionospheric**. The ionospheric path is important to medium-wave and HF propagation, but is not important to VHF, UHF, or microwave propagation. The space wave and surface wave are both **ground waves**, but behave differently. The surface wave travels in direct contact with the Earth's surface, and it suffers a severe frequency-dependent attenuation due to absorption into the ground.

The space wave is also a ground wave phenomenon, but is radiated from an antenna many wavelengths above the surface. No part of the space wave normally travels in contact with the surface; VHF, UHF, and microwave signals are usually space waves. There are, however, two components of the space wave in many cases: **direct** and **reflected** (Figure 1.2).

The ionosphere is the region of the Earth's atmosphere that is between the stratosphere and outer space. The peculiar feature of the ionosphere is that molecules of atmospheric gases (O_2 and N_2) can be ionized by stripping away electrons under the influence of solar radiation and certain other sources of energy (see Figure 1.1). In the ionosphere the air density is so low that positive ions can travel relatively long distances before recombining with electrons to form electrically neutral atoms. As a result, the ionosphere remains ionized for long periods of the day – even after sunset. At lower altitudes, however, air density is greater, and recombination thus occurs rapidly. At those altitudes, solar ionization diminishes to nearly zero imme-

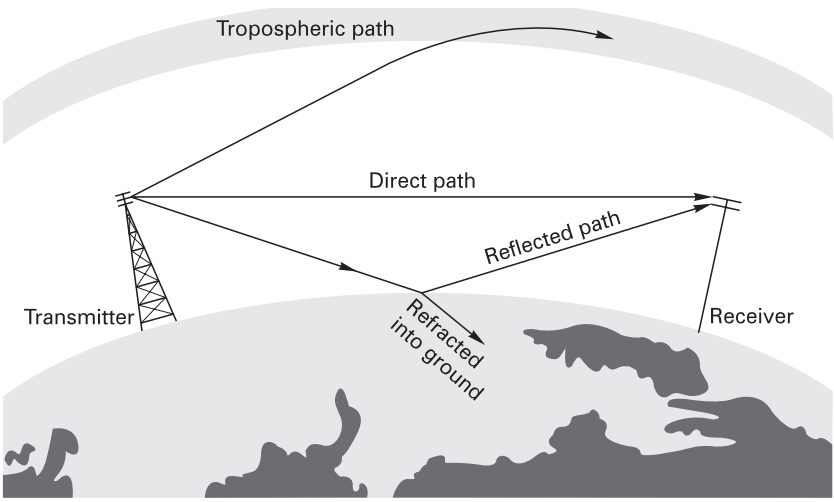


FIGURE 1.2

diately after sunset or never achieves any significant levels even at local noon.

Ionization and recombination phenomena in the ionosphere add to the noise level experienced at VHF, UHF, and microwave frequencies. The properties of the ionosphere are therefore important at these frequencies because of the noise contribution. In addition, in satellite communications there are some transionospheric effects.

GROUND WAVE PROPAGATION

The ground wave, naturally enough, travels along the ground, or at least in close proximity to it (Figure 1.3).

There are two basic forms of ground wave: **space wave** and **surface wave**. The space wave does not actually touch the ground. As a result, space wave attenuation with distance in clear weather is about the same as in free space (except above about 10 GHz, where absorption by H_2O and O_2 increases dramatically). Of course, above the VHF region, weather conditions add attenuation not found in outer space.

The surface wave is subject to the same attenuation factors as the space wave, but in addition it also suffers ground losses. These losses are due to ohmic resistive losses in the conductive earth. Surface wave attenuation is a function of frequency, and increases rapidly as frequency increases. For both of these forms of ground wave, communications is affected by the following factors: *wavelength*, *height of both receive and transmit antennas*, *distance between antennas*, and *terrain and weather along the transmission path*.

Ground wave communications also suffer another difficulty, especially at VHF, UHF, and microwave frequencies. The space wave is like a surface wave, but is radiated many wavelengths above the surface. It is made up of

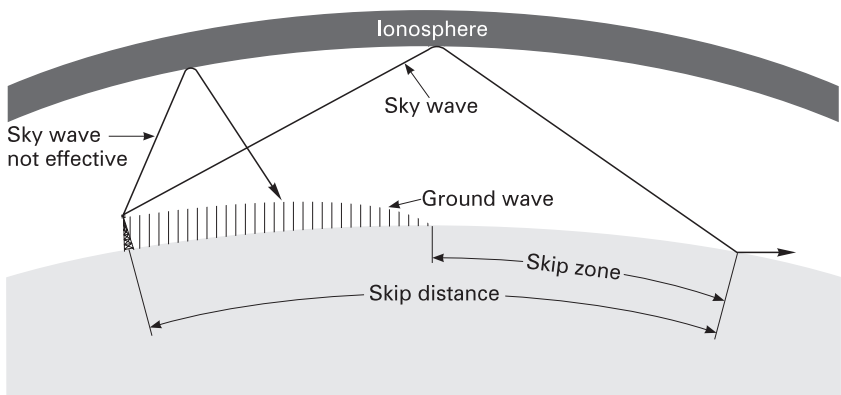


FIGURE 1.3

two components (see Figure 1.2): **direct** and **reflected** waves. If both of these components arrive at the receive antenna they will add algebraically to either increase or decrease signal strength. There is nearly always a phase shift between the two components because the two signal paths have different lengths. In addition, there may be a 180° (π radians) phase reversal at the point of reflection (especially if the incident signal is horizontally polarized).

Multipath phenomena exist because of interference between the direct and reflected components of the space wave. The form of multipath phenomenon that is, perhaps, most familiar to many readers (at least those old enough to be 'pre-cable') is ghosting in television reception. Some multipath events are transitory in nature (as when an aircraft flies through the transmission path), while others are permanent (as when a large building or hill reflects the signal). In mobile communications, multipath phenomena are responsible for reception dead zones and 'picket fencing.' A dead zone exists when destructive interference between direct and reflected (or multiple reflected) waves drastically reduces signal strengths. This problem is most often noticed at VHF and above when the vehicle is stopped; and the solution is to move the antenna a quarter wavelength. Picket fencing occurs as a mobile unit moves through successive dead zones and signal enhancement (or normal) zones, and sounds like a series of short noise bursts.

At VHF, UHF, and microwave frequencies the space wave is limited to so-called 'line of sight' distances. The horizon is theoretically the limit of communications distance, but the radio horizon is actually about 15% further than the optical horizon. This phenomenon is due to refractive bending in the atmosphere around the curvature of the Earth, and makes the geometry of the situation look as if the Earth's radius is four-thirds the actual radius.

The surface wave travels in direct contact with the Earth's surface, and it suffers a severe frequency-dependent attenuation due to absorption by the ground (Figure 1.3). The zone between the end of the ground wave and where the sky wave touches down is called the **skip zone**, and is a region of little or no signal. Because of this phenomenon, I have seen situations on the 15 m band (21.390 MHz) where two stations 65 km apart (Baltimore, Maryland, and Fairfax, Virginia) could not hear each other, and their communications have to be relayed via a ham station in Lima, Peru!

The surface wave extends to considerable heights above the ground level, although its intensity drops off rapidly at the upper end. The surface wave is subject to the same attenuation factors as the space wave, but in addition it also suffers ground losses. These losses are due to ohmic resistive losses in the conductive earth, and to the dielectric properties of the Earth. Horizontally polarized waves are not often used for surface wave communications because the Earth tends to short circuit the electrical (E) field

component. On vertically polarized waves, however, the Earth offers electrical resistance to the *E*-field and returns currents to following waves. The conductivity of the soil determines how much energy is returned.

IONOSPHERIC PROPAGATION

Now let us turn our attention to the phenomena of **skip communications** as seen in the short-wave bands, plus portions of the medium-wave and lower VHF regions. Ionospheric propagation is responsible for intercontinental broadcasting and communications.

Long-distance radio transmission is carried out on the HF bands (3–30 MHz), also called the 'short-wave' bands. These frequencies are used because of the phenomenon called **skip**. Under this type of propagation the Earth's ionosphere acts as if it is a 'radio mirror,' to reflect the signal back to Earth. This signal is called the **sky wave**. Although the actual phenomenon is based on refraction (not reflection, as is frequently believed) the appearance to the casual ground observer is that short-wave and low-VHF radio signals are reflected from the ionosphere as if it were a kind of radio mirror. The actual situation is a little different, but we will deal with that issue in a moment.

The key lies in the fact that a seeming radio mirror is produced by ionization of the upper atmosphere. The upper portion of the atmosphere is called the 'ionosphere' because it tends to be easily **ionized** by solar and cosmic radiation phenomena. The reason for the ease with which that region (50–500 km above the surface) ionizes is that the air density is very low. Energy from the Sun strips away electrons from the outer shells of oxygen and nitrogen molecules, forming free electrons and positive ions. Because the air is so rarified at those altitudes, these charged particles can travel great distances before recombining to form electrically neutral atoms again. As a result, the average ionization level remains high in that region.

Several sources of energy will cause ionization of the upper atomosphere. Cosmic radiation from outer space causes some degree of ionization, but the majority of ionization is caused by solar energy. The role of cosmic radiation was first noticed during World War II when military radar operators discovered that the distance at which their equipment could detect enemy aircraft was dependent upon whether or not the Milky Way was above the horizon (although it was theorized 10 years earlier). Intergalactic radiation raised the background microwave noise level, thereby adversely affecting the signal-to-noise ratio.

The ionosphere is divided for purposes of radio propagation studies into various layers that have different properties. These layers are only well defined in textbooks, however, and even there we find a variation in the height above the Earth's surface where these layers are found. In addition,

the real physical situation is such that layers do not have sharply defined boundaries, but rather fade one into another. The division into layers is therefore somewhat arbitrary. These layers (shown earlier in Figure 1.1) are designated D, E, and F (with F being further subdivided into the F_1 and F_2 sublayers).

D-layer

The D-layer is the lowest layer in the ionosphere, and exists from approximately 50 to 90 km above the Earth's surface. This layer is not ionized as much as higher layers because all forms of solar energy that cause ionization are severely attenuated by the higher layers above the D-layer. Another reason is that the D-layer is much denser than the E- and F-layers, and that density of air molecules allows ions and electrons to recombine to form electroneutral atoms very quickly.

The extent of D-layer ionization is roughly proportional to the height of the Sun above the horizon, so will achieve maximum intensity at midday. The D-layer exists mostly during the warmer months of the year because of both greater height of the sun above the horizon and the longer hours of daylight. The D-layer almost completely disappears after local sunset, although some observers have reported sporadic incidents of D-layer activity for a considerable time past sunset. The D-layer exhibits a large amount of absorption of medium-wave and short-wave signals, to such an extent that signals below 4–6 MHz are completely absorbed by the D-layer.

E-layer

The E-layer exists at altitudes between approximately 100 and 125 km. Instead of acting as an attenuator it acts primarily as a reflector although signals do undergo a degree of attenuation.

Like the D-layer, ionization in this region only exists during daylight hours, peaking around midday and falling rapidly after sunset. After nightfall the layer virtually disappears although there is some residual ionization there during the night-time hours.

The distance that is generally accepted to be maximum that can be achieved using E-layer propogation is 2500 km, although it is generally much less than this and can be as little as 200 km.

One interesting and exciting aspect of this region is a phenomenon called Es or sporadic E. When this occurs a layer or cloud of very intense ionization forms. This can reflect signals well into the VHH region of the radio spectrum. Although generally short lived, there can be openings on bands as high as 2 meters (144 MHz). These may last as little as a few minutes, whilst long openings may last up to a couple of hours. The phenomenon also affects lower frequencies like the 10 meter and 6 meter amateur bands as

well as the VHF FM band. Sporadic E is most common in the summer months, peaking in June (in the northern hemisphere). Distances of between 1000 and 2500 km can be reached using this mode of propagation.

F-layer

The F-layer of the ionosphere is the region that is the principal cause of long-distance short-wave communications. This layer is located from about 150–500 km above the Earth's surface. Unlike the lower layers, the air density in the F-layer is low enough that ionization levels remain high all day, and decay slowly after local sunset. Minimum levels are reached just prior to local sunrise. Propagation in the F-layer is capable of skip distances up to 4000 km in a single hop. During the day there are actually two identifiable, distinct sublayers in the F-layer region, and these are designated the 'F_1' and 'F_2' layers. The F_1 layer is found approximately 150–250 km above the Earth's surface, while the F_2 layer is above the F_1 to the 450–500 km limit. Beginning at local sundown, however, the lower regions of the F_1 layer begin to de-ionize due to recombination of positive ions and free electrons. At some time after local sunset the F_1 and F_2 layers have effectively merged to become a single reduced layer beginning at about 300 km.

The height and degree of ionization of the F_2 layer varies over the course of the day, with the season of the year, and with the 27 day cycle of the sun. The F_2 layer begins to form shortly after local sunrise, and reaches a maximum shortly before noon. During the afternoon the F_2 layer ionization begins to decay in an exponential manner until, for purposes of radio propagation, it disappears sometime after local sunset. There is some evidence that ionization in the F-layer does not completely disappear, but its importance to HF radio communication does disappear.

IONOSPHERIC VARIATION AND DISTURBANCES

The ionosphere is an extremely dynamic region of the atmosphere, especially from a radio operator's point of view, for it significantly alters radio propagation. The dynamics of the ionosphere are conveniently divided into two general classes: **regular variation** and **disturbances**. We will now look at both types of ionospheric change.

Ionospheric variation

There are several different forms of variation seen on a regular basis in the ionosphere: **diurnal**, **27 day** (monthly), **seasonal**, and **11 year cycle**.

Diurnal (daily) variation

The Sun rises and falls in a 24 hour cycle, and because it is a principal source of ionization of the upper atmosphere, one can expect diurnal variation. During daylight hours the E- and D-levels exist, but these disappear at night. The height of the F_2 layer increases until midday, and then decreases until evening, when it disappears or merges with other layers. As a result of higher absorption in the E- and D-layers, lower frequencies are not useful during daylight hours. On the other hand, the F-layers reflect higher frequencies during the day. In the 1–30 MHz region, higher frequencies (>11 MHz) are used during daylight hours and lower frequencies (<11 MHz) at night. Figure 1.4B shows the number of sunspots per year since 1700.

27 day cycle

Approximately monthly in duration, this variation is due to the rotational period of the Sun. Sunspots (Figure 1.4A) are localized on the surface of the Sun, so will face the Earth only during a portion of the month. As new sunspots are formed, they do not show up on the earthside face until their region of the Sun rotates earthside.

Seasonal cycle

The Earth's tilt varies the exposure of the planet to the Sun on a seasonal basis. In addition, the Earth's yearly orbit is not circular, but elliptical. As a result, the intensity of the Sun's energy that ionizes the upper atmosphere varies with the seasons of the year. In general, the E-, D-, and F-layers are affected, although the F_2 layer is only minimally affected. Ion density in the F_2 layer tends to be highest in winter, and less in summer. During the summer, the distinction between F_1 and F_2 layers is less obvious.

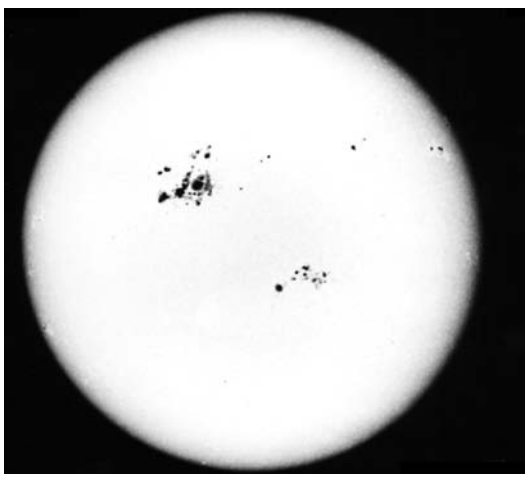


FIGURE 1.4A

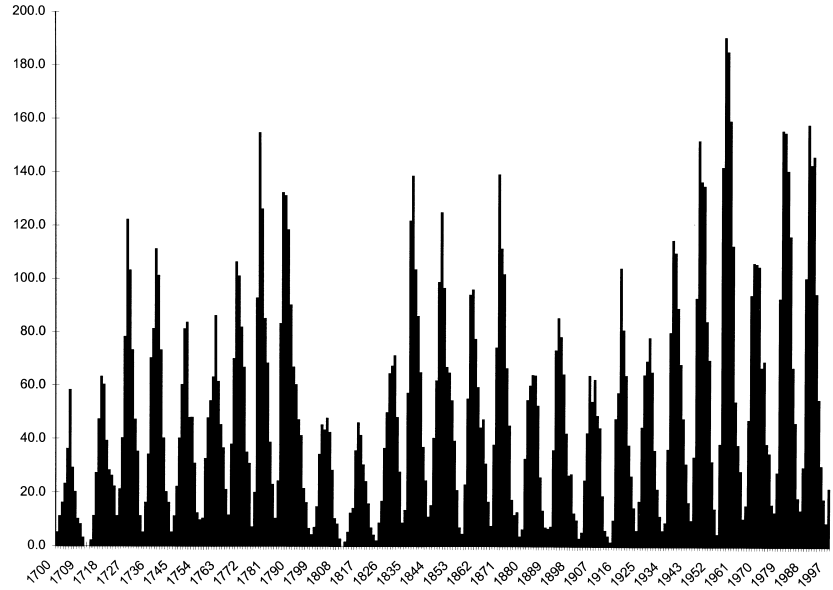


FIGURE 1.4B

11 year cycle

The number of sunspots, statistically averaged, varies on an approximately 11 year cycle (Fig. 1.4B). As a result, the ionospheric effects that affect radio propagation also vary on an 11 year cycle. Radio propagation in the short-wave bands is best when the average number of sunspots is highest. Peaks occurred in 1957, 1968, 1979, and 1990.

Events on the surface of the Sun sometimes cause the radio mirror to seem almost perfect, and make spectacular propagation possible. At other times, however, solar disturbances disrupt radio communications for days at a time.

There are two principal forms of solar energy that affect short-wave communications: **electromagnetic radiation** and **charged solar particles**. Most of the radiation is beyond the visible spectrum, in the ultraviolet and X-ray/γ-ray region of the spectrum. Because electromagnetic radiation travels at the speed of light, solar events that release radiation cause changes to the ionosphere about 8 minutes later. Charged particles, on the other hand, have a finite mass and so travel at a considerably slower velocity. They require 2 or 3 days to reach the Earth.

Various sources of both radiation and particles exist on the Sun. Solar flares may release huge amounts of both radiation and particles. These events are unpredictable and sporadic. Solar radiation also varies over an approximately 27 day period, which is the rotational period of the Sun. The same source of radiation will face the Earth once every 27 days, so events tend to be somewhat repetitive.

Solar and galactic noise affect the reception of weak signals, while solar noise will also either affect radio propagation or act as a harbinger of changes in propagation patterns. Solar noise can be demonstrated by using an ordinary radio receiver and a directional antenna, preferably operating in the VHF/UHF regions of the spectrum. If the antenna is aimed at the Sun on the horizon at either sunset or sunrise a dramatic change in background noise will be noted as the Sun slides across the horizon.

Sunspots

A principal source of solar radiation, especially the periodic forms, is sunspots (Figure 1.4A). Sunspots can be as large as 100 000–150 000 km in diameter, and generally occur in clusters. The number of sunspots varies over a period of approximately 11 years, although the actual periods since 1750 (when records were first kept) have varied from 9 to 14 years (Fig. 1.4B). The sunspot number is reported daily as the statistically massaged **Zurich smoothed sunspot number**, or **Wolf number**. The number of sunspots greatly affects radio propagation via the ionosphere. The low was in the range of 60 (in 1907), while the high was about 200 (1958).

Another indicator of ionospheric propagation potential is the **solar flux index** (SFI). This measure is taken in the microwave region (wavelength of 10.2 cm, or 2.8 GHz), at 1700 U.T. Greenwich Mean Time in Ottawa, Canada. The SFI is reported by the National Institutes of Standards and Technology (NIST) radio stations WWV (Fort Collins, Colorado) and WWVH (Maui, Hawaii).

The ionosphere offers different properties that affect radio propagation at different times. Variations occur not only over the 11 year sunspot cycle but also diurnally and seasonally. Obviously, if the Sun affects propagation in a significant way, then differences between night-time and daytime, and between summer and winter, must cause variations in the propagation phenomena observed.

Ionospheric disturbances

Disturbances in the ionosphere can have a profound effect on radio communications – and most of them (but not all) are bad. In this section we will briefly examine some of the more common forms.

Sporadic E-layer

A reflective cloud of ionization sometimes appears in the E-layer of the ionosphere; this layer is sometimes called the E_s layer. It is believed that the E_s layer forms from the effects of wind shear between masses of air moving in opposite directions. This action appears to redistribute ions into a thin a layer that is radio-reflective.

Sporadic-E propagation is normally thought of as a VHF phenomenon, with most activity between 30 and 100 MHz, and decreasing activity up to about 100 MHz. However, about 25–50% of the time, sporadic-E propagation is possible on frequencies down to 10–15 MHz. Reception over paths of 2300–4200 km is possible in the 50 MHz region when sporadic-E propagation is present. In the northern hemisphere, the months of June and July are the most prevalent sporadic-E months. On most days when the sporadic-E phenomenon is present it lasts only a few hours.

Sudden ionospheric disturbances (SIDs)

The SID, or 'Dellinger fade,' mechanism occurs suddenly, and rarely gives any warning. Solar flares (Figure 1.5) are implicated in SIDs. The SID may last from a few minutes to many hours. It is believed that SIDs occur in correlation with solar flares or 'bright solar eruptions' that produce immense amounts of ultraviolet radiation that impinge the upper atmosphere. The SID causes a tremendous increase in D-layer ionization, which accounts for the radio propagation effects. The ionization is so intense that all receiver operators on the sunny side of the Earth experience profound loss of signal strength above about 3 MHz. It is not uncommon for receiver owners to think that their receivers are malfunctioning when this

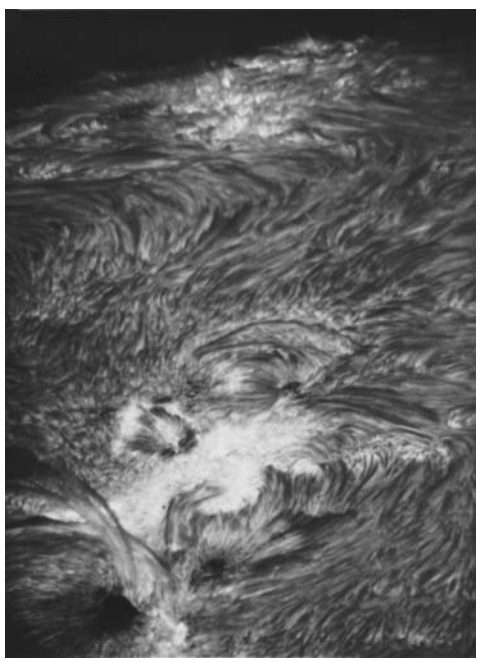


FIGURE 1.5

occurs. The sudden loss of signal by sunny-side receivers is called **Dellinger fade**. The SID is often accompanied by variations in terrestial electrical currents and magnetism levels.

An interesting anomaly is seen when SIDs occur. Although short-wave reception is disrupted, and may stay that way for awhile, distant very low-frequency (VLF) signals, especially in the 15–40 kHz region, experience a sudden increase in intensity. This is due to the fact that the SID event results in deep ionization way into the D-layer. This ionization increases absorption of HF signals. But the wavelength of VLF signals is close to the distance from the Earth's surface to the bottom of the D-layer, so that space acts like a gigantic 'waveguide' (as used in the transmission of microwaves) when the SID is present – thus propagating the VLF signal very efficiently.

Ionospheric storms

The ionospheric storm appears to be produced by an abnormally large rain of atomic particles in the upper atmosphere, and is often preceded by a SID 18–24 hours earlier. These storms tend to last from several hours to a week or more, and are often preceded by 2 days or so by an abnormally large collection of sunspots crossing the solar disk. They occur most frequently, and with greatest severity, in the higher latitudes, decreasing toward the Equator. When the ionospheric storm commences, short-wave radio signals may begin to flutter rapidly and then drop out altogether. The upper ionosphere becomes chaotic, turbulence increases, and the normal stratification into 'layers' or zones diminishes.

Radio propagation may come and go over the course of the storm, but it is mostly absent. The ionospheric storm, unlike the SID which affects the sunny side of the Earth, is worldwide. It is noted that the maximum usable frequency (MUF) and critical frequency tend to reduce rapidly as the storm commences.

An ionospheric disturbance observed over November 12–14, 1960 was preceded by about 30 minutes of extremely good, but abnormal propagation. At 15.00 hours EST, European stations were noted in North America with S9+ signal strengths in the 7000–7300 kHz region of the spectrum, which is an extremely rare occurrence. After about 30 minutes, the bottom dropped out, and even AM broadcast band skip (later that evening) was non-existent. At the time, WWV was broadcasting a 'W2' propagation prediction at 19 and 49 minutes after each hour. It was difficult to hear even the 5 MHz WWV frequency in the early hours of the disturbance, and it disappeared altogether for the next 48 hours. Of course, as luck would have it, that event occurred during the first weekend of the ARRL Sweepstakes ham radio operating contest that year.

GREAT CIRCLE PATHS

A great circle is a line between two points on the surface of a sphere that lies on a plane through the sphere's center. When translated to 'radio speak,' a great circle is the shortest path on the surface of the Earth between two points. Navigators and radio operators use the great circle for similar, but different reasons: the navigator in order to get from here to there, and the radio operator to get a transmission path from here to there.

The heading of a directional antenna is normally aimed at the receiving station along its great circle path. Unfortunately, many people do not understand the concept well enough, for they typically aim the antenna in the wrong direction. Radio waves do not travel along what appears to be the best route on a flat map. Instead they travel along the shortest distance on a real globe.

Long path versus short path

The Earth is a sphere (or more precisely, an 'oblique spheroid'), so from any given point to any other point there are two great circle paths: the **long path** (major arc) and **short path** (minor arc). In general, the best reception occurs along the short path. In addition, short-path propagation is more nearly 'textbook' compared with long-path reception. However, there are times when the long path is better, or is the only path that will deliver a signal to a specific location from the geographic location in question.

USING THE IONOSPHERE

The refraction of HF and some medium-wave radio signals back to Earth via the ionosphere gives rise to intercontinental HF radio communications. This phenomenon becomes possible during daylight hours, and for a while after sunset when the ionosphere is ionized. Figure 1.6 reiterates the mechanism of long-distance skip communications. The transmitter is located at point T, while receiving stations are located at sites R1 and R2. Signals 1 and 2 are not refracted sufficiently to be returned to Earth, so they are lost in space. Signal 3, however, is refracted enough to return to Earth, so it is heard at station Rl. The skip distance for signal 3 is the distance from T to Rl. At points between T and R1, signal 3 is inaudible, except within ground wave distance of the transmitter site (T). This is the reason why two stations 50 km apart hear each other only weakly, or not at all, while both stations can communicate with a third station 3000 km away. In American amateur radio circles it is common for South American stations to relay between two US stations only a few kilometers apart.

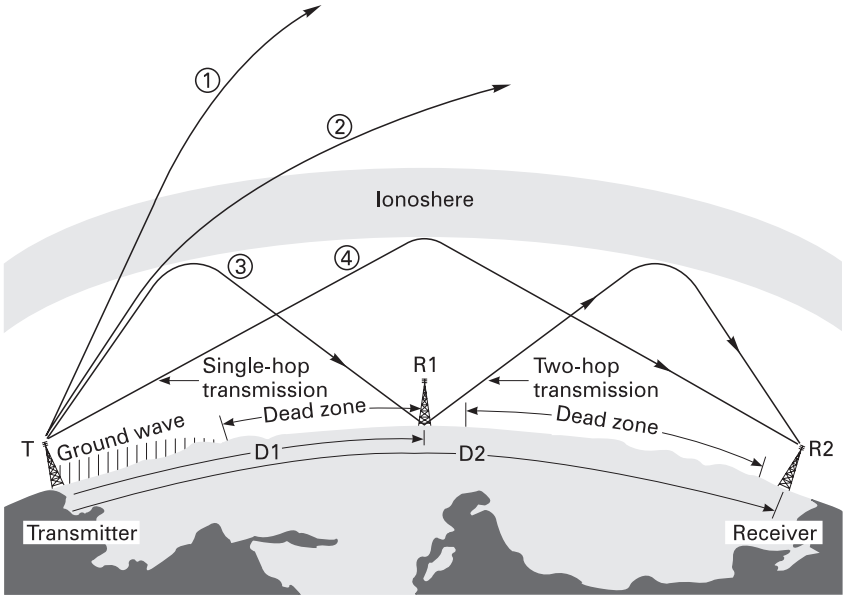


FIGURE 1.6

Multi-hop skip is responsible for the reception of signal 3 at site R2. The signal *reflects* (not refracts) from the surface at R1, and is retransmitted into the ionosphere, where it is again refracted back to Earth.

The location where skip signals are received (at different distances) depends partially upon the **angle of radiation** of the transmitting antenna. A high angle of radiation causes a shorter skip zone, while a lower angle of radiation results in a longer skip zone. Communication between any particular locations on any given frequency requires adjustment of the antenna radiation angle. Some international short-wave stations have multiple antennas with different radiation angles to ensure that the correct skip distances are available.

SUPER-REFRACTION AND SUBREFRACTION

At VHF frequencies and well up into the microwave bands there are special propagation modes called *super-refraction* and *subrefraction*. Depending upon the temperature gradient and humidity, the propagation may not be straight line. At issue is something called the *K-factor* of the wave. Figure 1.7 shows these modalities. The straight line case has a K-factor of one, and is the reference. If super-refraction occurs, then the value of K is greater than one, and if subrefraction occurs it is less than one.

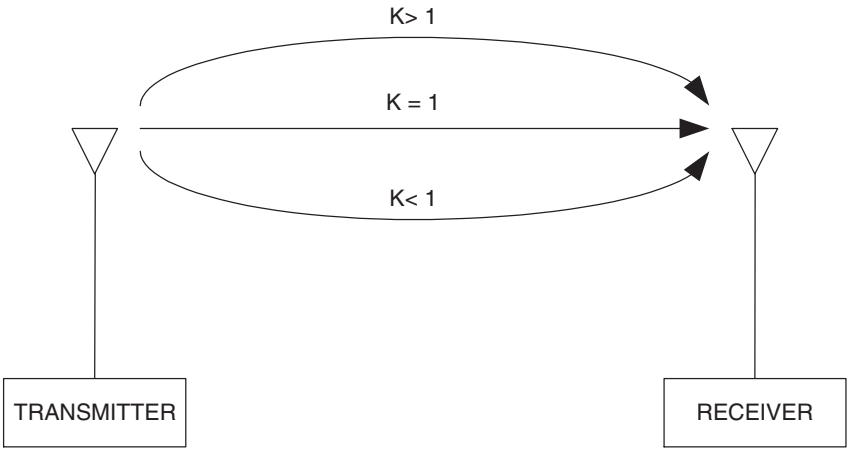


FIGURE 1.7

Figure 1.8 shows a case where super-refraction occurs. The value of K can be substantially above one in cases where a hot body of land occurs next to a relatively cool body of water. This occurs off Baja, California, the Persian Gulf and Indian Ocean, parts of Australia and the North African coast. In those areas, there may be substantial amounts of super-refraction occurring, making directional antennas point in the wrong direction.

Figure 1.9 shows a case of subrefraction. In this case, there is a relatively cold land mass next to relatively warm seas. In the Arctic and Antarctic this situation exists. The K factor will be substantially less than one in these cases. In fact, the signal may be lost to terrestrial communications after only a relatively short distance.

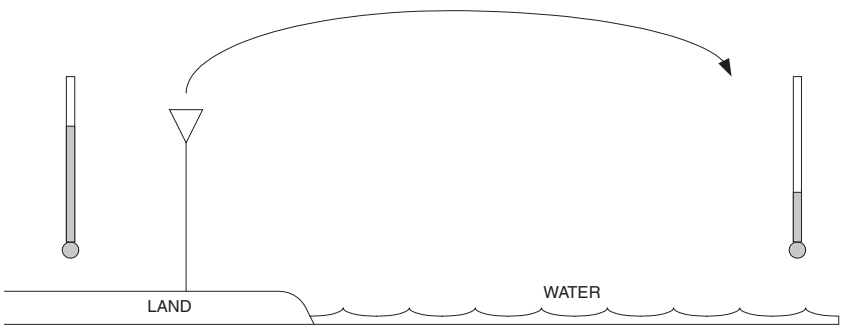


FIGURE 1.8

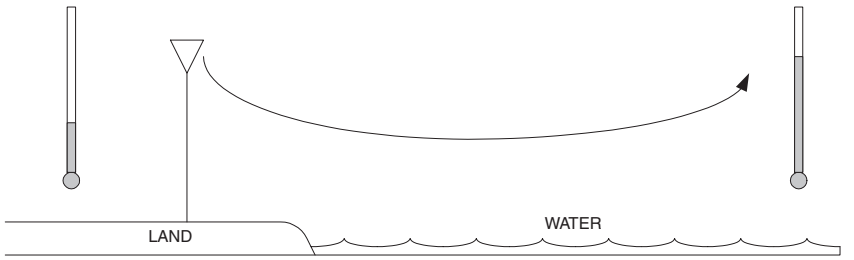


FIGURE 1.9

CHAPTER 2

Antenna basics

Before looking at the various antennas we need to look at some of the basics of antenna systems. In this chapter you will learn some of these basics. And while they will not make you a red-hot professional antenna engineer they will set you up well enough to understand this book and others on amateur and hobbyist antennas. We will look at the matter of antenna radiation, antenna patterns, the symbols used to represent antennas, voltage standing wave ratio (VSWR), impedance, and various methods suitable for constructing wire antennas in the high-frequency (HF) and very high-frequency (VHF) regions of the spectrum.

ANTENNA SYSTEM SYMBOLS

Figures 2.1 and 2.2 show the various symbols used to represent antennas and grounds. The reason why there are so many variants is that there are differences from country to country, as well as different practices within any one country (especially between technical publishers). As for antenna symbols, I see the symbol in Figure 2.1C more often in the USA, but Figure 2.1B comes in for a close second. The supposedly correct symbol (endorsed by a professional society drawing standards committee) is that of Figure 2.1A – but it is only occasionally seen in the USA.

The situation for grounds is a little different because some differences reflect different forms of ground (although some of the differences also represent national or publisher differences). The ground in Figure 2.2A is usually found representing a true earth ground, i.e. the wire is connected to a

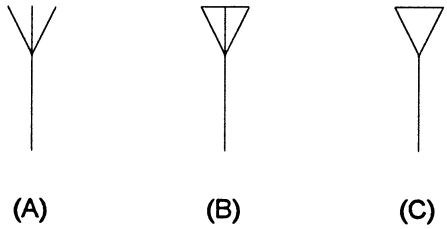


FIGURE 2.1

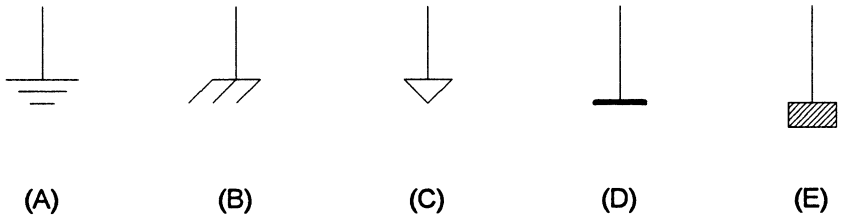


FIGURE 2.2

rod driven into the earth. The variant of Figure 2.2B usually represents a chassis ground inside a piece of equipment. The symbol in Figure 2.2C has two uses. One is to represent a common grounding point for different signals or different pieces of equipment. The other use is exactly the opposite: the triangle ground symbol often represents an **isolated ground** that has no direct electrical connection to the rest of the circuit, or with the earth. You will see this usage in medical devices. The grounds of Figures 2.2D and 2.2E are found mostly outside the USA.

From the time of Hertz and Marconi to the present, one thing has remained constant in wireless communications: radio waves travel, as if by magic, from a transmitting antenna to a receiving antenna (Figure 2.3). Whether the two antennas are across the garden from each other, across continents and oceans, or on the Earth and the Moon, if there is not a transmitting antenna and at least one receiving antenna in the system then no communications can take place.

At one time, physicists believed that there must be some invisible medium for carrying the radio signal. But we now know that no such medium exists, yet radio waves travel even in outer space. Being electromagnetic waves, radio signals need no medium in order to propagate. If radio signals traveled only in the Earth's atmosphere, then we could make some guesses about a medium for carrying the wave, but space communications demonstrates that the atmosphere is not necessary (although it does affect radio signal propagation).

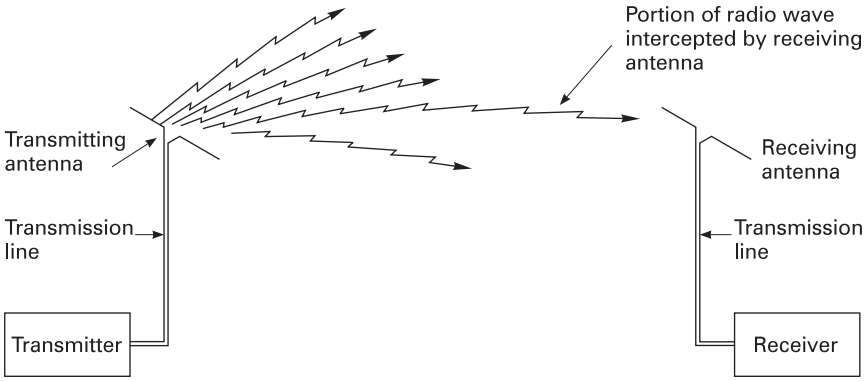
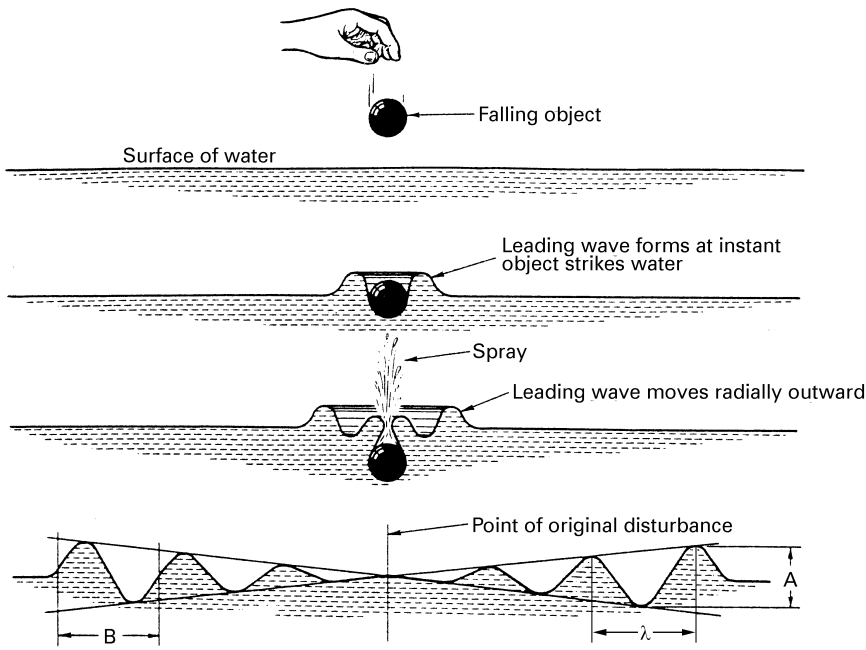


FIGURE 2.3

Although there is no medium in which radio waves travel, it is useful to look at water waves for an analogy (even though imperfect). In Figure 2.4 we see what happens when an object is dropped into a pool of water. A displacement takes place, which forms a leading wave that pushes out in concentric circles from the impact point. The situation in Figure 2.4 repre-



Notes: A, amplitude of leading wave.
B, corresponds to 1 cycle of oscillation.

FIGURE 2.4

sents a single pulse of energy, as if a transmitter fired a single burst of energy. Real transmitters send out wave trains that are analogous to cyclically bobbing the object up and down so that it goes in and out of the water (Figure 2.5). The result is a continuous stream of identical waves propagating out from the 'transmitter' impact point. If another object is floating on the surface, say a cork or toy boat, then it will be perturbed as the wave passes. This is analogous to the receiver antenna.

The waves have an **amplitude** ('A'), which corresponds to the signal strength. They also have a **wavelength** (λ), which corresponds to the distance traveled by the wave in one complete up-and-down cycle. In radio work, the wavelength is measured in **meters** (m), except in the microwave region where **centimeters** (cm) and **millimeters** (mm) make more sense. Wavelength can be measured at any pair of points on the wave that are identical: two peaks, two troughs, two zero crossings, as convenient in any specific case.

The number of cycles that pass a given point every second is the **frequency** of the wave. The classic measure of frequency was cycles per second (cps or c/s), but that was changed in 1960 by international consensus to the hertz (Hz), in honor of Heinrich Hertz. But since 1 Hz = 1 cps, there is no

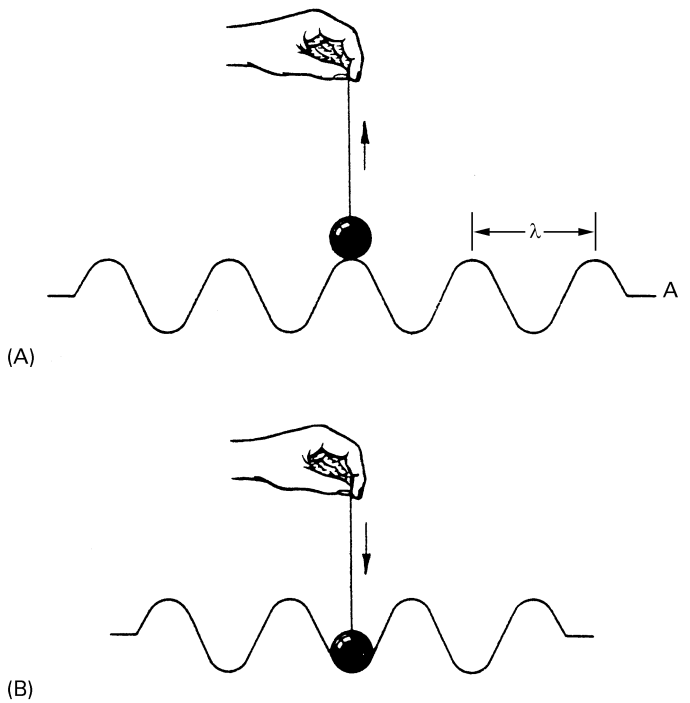


FIGURE 2.5

practical difference. The hertz is too small a unit for most radio work (although many of our equations are written in terms of hertz). For radio work the **kilohertz** (kHz) and **megahertz** (MHz) are used: 1 kHz = 1000 Hz, and 1 MHz = 1 000 000 Hz. Thus, a short-wave frequency of 9.75 MHz is 9750 kHz and 9 750 000 Hz.

Wavelength and frequency are related to each other. The wavelength is the reciprocal of frequency, and vice versa, through the velocity constant. In free space, the velocity constant is the speed of light (c), or about 300 000 000 m/s. This is the reason why you often see '300' or its submultiples (150 and 75) in equations. When the frequency is specified in megahertz, then 300 000 000 becomes 300 for one wavelength. The half-wavelength constant is 150, and the quarter-wavelength constant is 75. The relationship is

$$\lambda_{\text{meters}} = \frac{300}{F_{\text{MHz}}}$$

INVERSE SQUARE LAW

When radio waves travel they become weaker by a relationship called the **inverse square law**. This means that the strength is inversely proportional to the square of the distance traveled ($1/D^2$). Figure 2.6 shows how this works using the analogy of a candle. If the candle projects a distance r, all of the light energy falls onto square 'A'. At twice the distance ($2r$) the light spreads out and covers four times the area (square 'B'). The total amount of light energy is the same, but the energy per unit of area is reduced to one-fourth of the energy that was measured at 'A.' This means that a radio signal gets weaker very rapidly as the distance from the transmitter increases, requiring ever more sensitive receivers and better antennas.

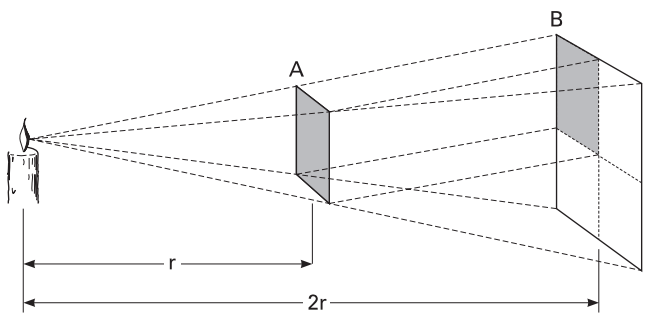


FIGURE 2.6

THE ELECTROMAGNETIC WAVE

The electromagnetic (EM) wave propagating in space is what we know as a 'radio signal.' The EM wave is launched when an electrical current oscillates in the transmitting antenna (Figure 2.7). Because moving electrical currents possess both electrical (*E*) and magnetic (*H*) fields, the electromagnetic wave launched into space has alternating *E*-field and *H*-field components. These fields are **transverse** (meaning they travel in the same direction) and **orthogonal** (meaning the *E*- and *H*-fields are at right angles to each other). When the EM wave intercepts the receiver antenna, it sets up a copy of the original oscillating currents in the antenna, and these currents are what the receiver circuitry senses.

The orthogonal *E*- and *H*-fields are important to the antenna designer. If you could look directly at an oncoming EM wave, you would see a plane front advancing from the transmitting antenna. If you had some magical dye that would render the *E*-field and *H*-field line of force vectors visible to the naked eye, then you would see the *E*-field pointing in one direction, and the *H*-field in a direction 90° away (Figure 2.8).

The **polarization** of the signal is the *direction of the* E-*field vector*. In Figure 2.8 the polarization is vertical because the electric field vector is up and down. If the *E*-field vector were side-to-side, then the polarization would be horizontal. One way to tell which polarization an antenna produces when it transmits, or is most sensitive to when it receives, is to note the direction of the radiator element. If the radiator element is vertical, i.e. perpendicular to the Earth's surface, then it is vertically polarized. But if

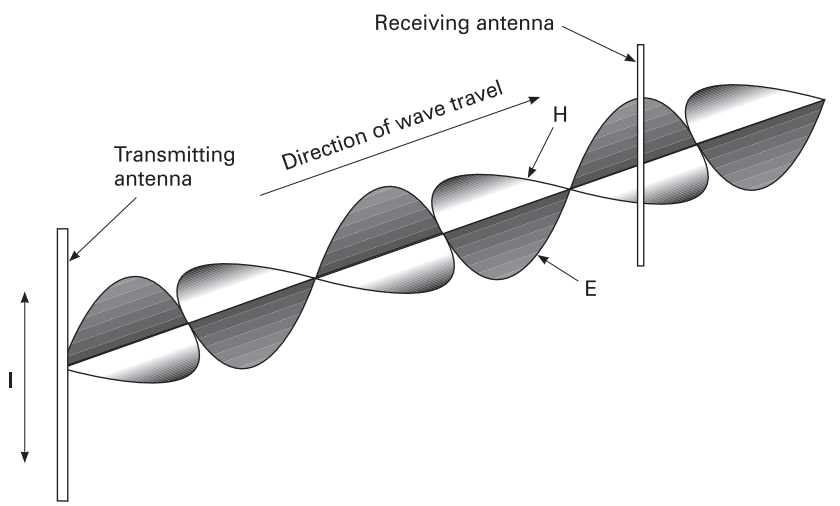


FIGURE 2.7

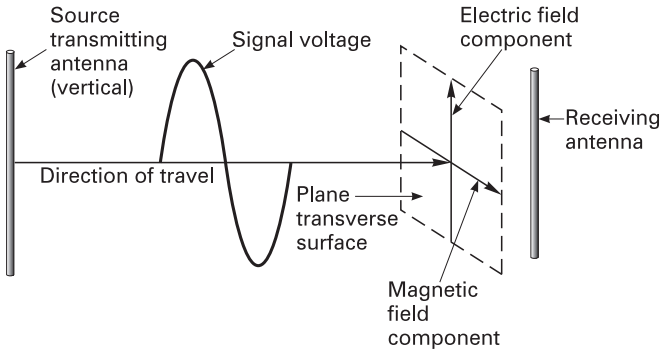


FIGURE 2.8

the radiator element is horizontal with respect to the Earth's surface, then it is horizontally polarized. Figure 2.9 shows these relationships. In Figure 2.9, two dipole receiver antennas are shown, one is vertically polarized (VD) and the other is horizontally polarized (HD). In Figure 2.9A, the arriving signal is vertically polarized. Because the *E*-field vectors lines are vertical, they cut across more of the VD antenna than the HD, producing a considerably larger signal level. The opposite is seen in Figure 2.9B. Here the *E*-field is horizontally polarized, so it is the HD antenna that receives the most signal. The signal level difference can be as much as 20 dB, which represents a 10-fold decrease in signal strength if the wrong antenna is used.

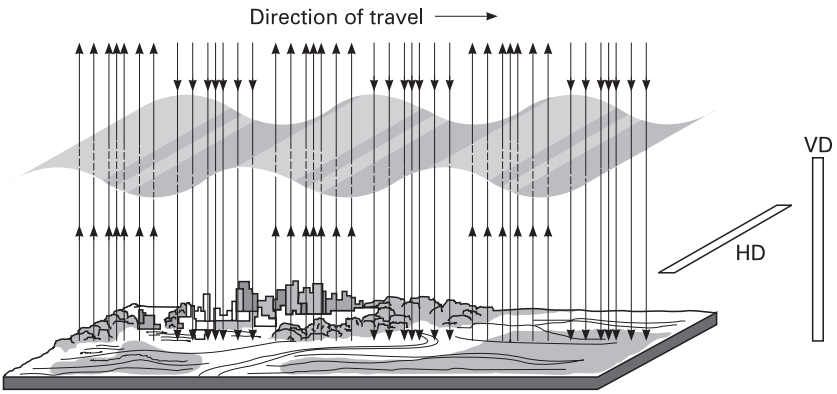
DECIBELS (dB)

In the section above the term **decibel** (symbol 'dB') was introduced. The decibel is a unit of measure of the ratio of two signals: two voltages, two currents, or two powers. The equations for decibels take the logarithm of the ratio, and multiply it by a constant (10 for powers and 20 for voltages or currents). The use of decibel notation makes it possible to use **ratios**, such as found in gains and losses in electronic circuits, but use only addition and subtraction arithmetic. It is not necessary to be able to calculate decibels, but you should know that +dB represents a gain, and −dB represents a loss. The term '0 dB' means that the ratio of the two signals is 1:1 (neither gain nor loss). Some common ratios encountered in radio work include those listed in Table 2.1.

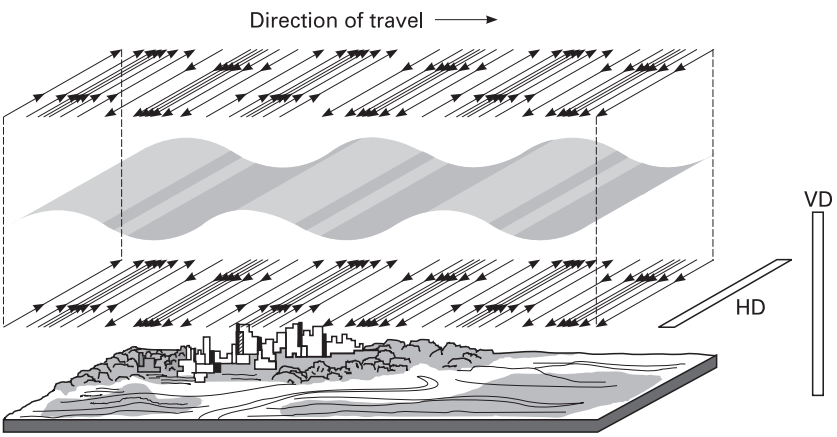
You can see that doubling a signal strength results in a +3 dB gain, while halving it produces a −3 dB loss. To put these figures into perspective, most S-meters on receivers use a scaling factor of 6 dB per S-unit (some use 3 dB per S-unit). Receiver designers tell us that a signal-to-noise ratio of 10 dB is necessary for 'comfortable listening,' while a signal-to-noise ratio of 3 dB is

TABLE 2.1

Ratio	Power (dB)	Voltage (dB)
1:1	0	0
2:1	+3	+6
1:2	−3	−6
10:1	+10	+20
1:10	−10	−20



(A) Vertical polarization



(B) Horizontal polarization

FIGURE 2.9

needed for barely perceptible but reliable communication for a listener who tries hard to hear what is being said.

LAW OF RECIPROCITY

Radio antennas obey a kind of **law of reciprocity**, i.e. they work the same on transmit as they do on receive. If an antenna has a certain gain and directivity on transmit, then that exact same pattern is seen in the receive mode. Similarly, the feedpoint impedance, element lengths, spacings, and other issues are the same for both modes. An implication of this law is that receiver owners are able to translate the theory from discussions of transmitting antennas to their own needs (and vice versa).

Reciprocity does not necessarily mean that the same antennas are the best selection for both transmit and receive. For example, the use of gain on a transmit antenna is to increase the apparent signal power level at a distant point in a particular direction. Alternatively, some licensing authorities use antenna directivity to protect the coverage areas of relatively nearby stations that share the same or adjacent frequencies. To the receiver operator, there may be a reason to want the gain of the antenna to boost the received signal power to a useful level. There is, however, often a powerful argument for aiming an antenna such that the main sensitivity is not in the direction of the desired station. Instead, the receiver antenna owner may wish to position a null, i.e. the least sensitive aspect of the antenna, in the direction of an offending station in order to reduce its effect. The issue is, after all is said and done, the signal-to-noise ratio.

Another restriction is that there are several antennas that are fit for receive use, especially those for difficult installation situations, but are either technically unsuited or unsafe for transmitters except at the lowest of power levels. The issue might be impedance, VSWR, voltage arcing, or some other undesirable factor that occurs in transmit situations (even at moderate power levels such as 50–100 W).

ANTENNA TYPES

Most antennas fall into either of two broad categories: **Marconi** or **Hertzian**. The Marconi antennas (Figure 2.10) are considered unbalanced with respect to ground because one side of the signal source (or receiver input load) is grounded. The Marconi antenna can be either vertical (Figure 2.10A) or at some angle (Figure 2.10B), including horizontal. The name derives from the types of antenna used by Guglielmo Marconi in his early radio experiments. In his famous 1903 transatlantic communication, ol' 'Gug' used a single wire tethered to a kite aloft above the Newfoundland coast.

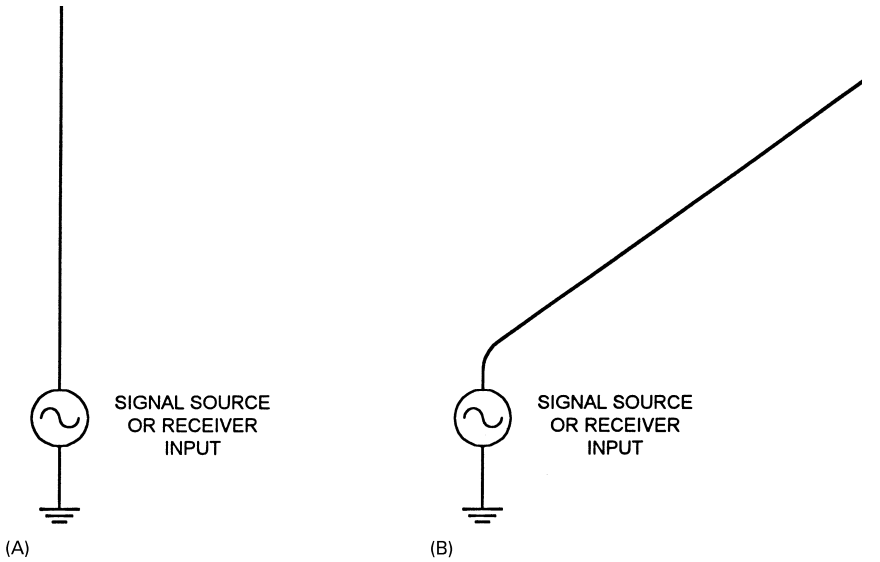


FIGURE 2.10

The Hertzian antennas (Figure 2.11) are balanced with respect to the ground, i.e. neither side of the oscillator or receiver load is grounded. Dipoles fall into this category. Both Marconi and Hertzian antennas can be either vertically or horizontally polarized. However, most Marconi antennas are designed for vertical polarization and most Hertzian antennas for horizontal polarization. The long wire and vertical dipoles are obvious exceptions to that rule, however.

RECEIVER–ANTENNA INTERACTIONS

The antenna and the receiver (or the antenna and transmitter on the other end) form a system that must be used together. Either alone is not too useful. Figure 2.12 shows an antenna connected to a receiver through a transmission line. For our practical purpose here, the receiver looks to the

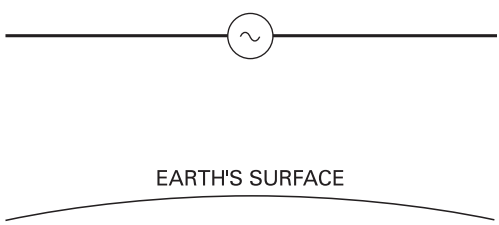


FIGURE 2.11

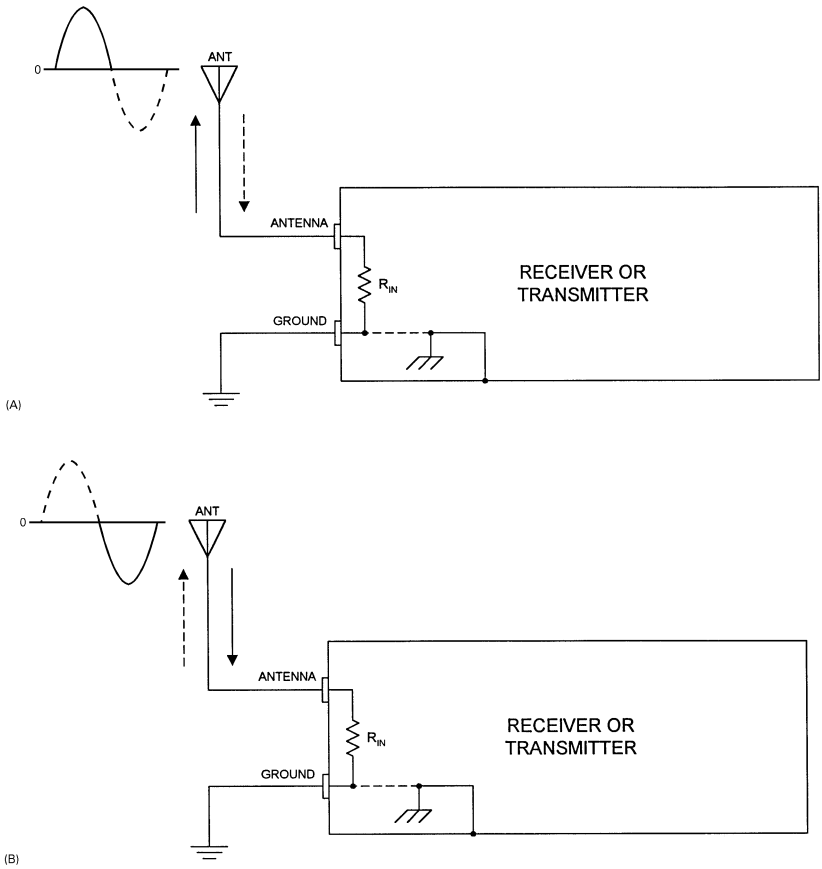


FIGURE 2.12

antenna and transmission line like a load resistor, R_{IN}. On the positive half-cycle (Figure 2.12A), the passing wave creates a current in the antenna–receiver system that flows in one direction (shown here as 'up'). When the oscillation reverses and becomes negative, the direction of current flow in the antenna–receiver reverses (Figure 2.12B). This is the mechanism by which the oscillating electromagnetic wave reproduces a signal in the input of the receiver – the signal that is then amplified and demodulated to recover whatever passes for 'intelligence' riding on the signal.

STANDING WAVE RATIO

The issue of the **standing wave ratio** (SWR) is of constant interest to radio enthusiasts. Some of the heat and smoke on this matter is well justified. In other cases, the perceived problems are not real.

Figure 2.13 shows how the SWR comes into play in an antenna system. In Figure 2.13A, a single cycle of a signal is launched down a transmission line (it is called the '**incident**' or '**forward**' wave). When it reaches the end of the line, if it is not totally absorbed by a load resistor or antenna, then it (or part of it) will be reflected back toward the source. This **reflected** wave is shown in Figure 2.13B. The incident and reflected waves are both examples of **travelling waves**. The reflected wave represents power that is lost, and can cause other problems as well.

The situation in Figures 2.13A and 2.13B represent a single-cycle pulse launched down a transmission line. In a real radio system, the oscillations of the incident wave are constant (Figure 2.13C). When this situation occurs, then the reflected waves will interfere with following incident waves. At any given point, the amplitude of the wave is the algebraic sum of the interfering incident and reflected signals. The resultant caused by the interference of the incident and reflected waves is called a **standing wave**.

Figure 2.14 shows what happens when continuous incident and reflected waves coexist on the same transmission line. In the case of Figure 2.14A, the two waves coincide, with the resultant as shown. The waves begin to move

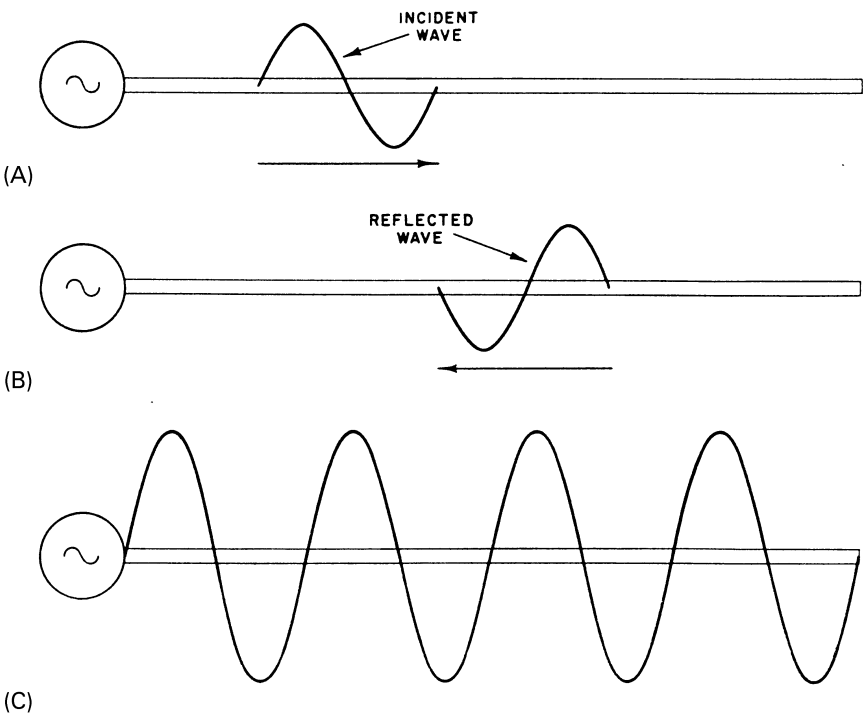


FIGURE 2.13

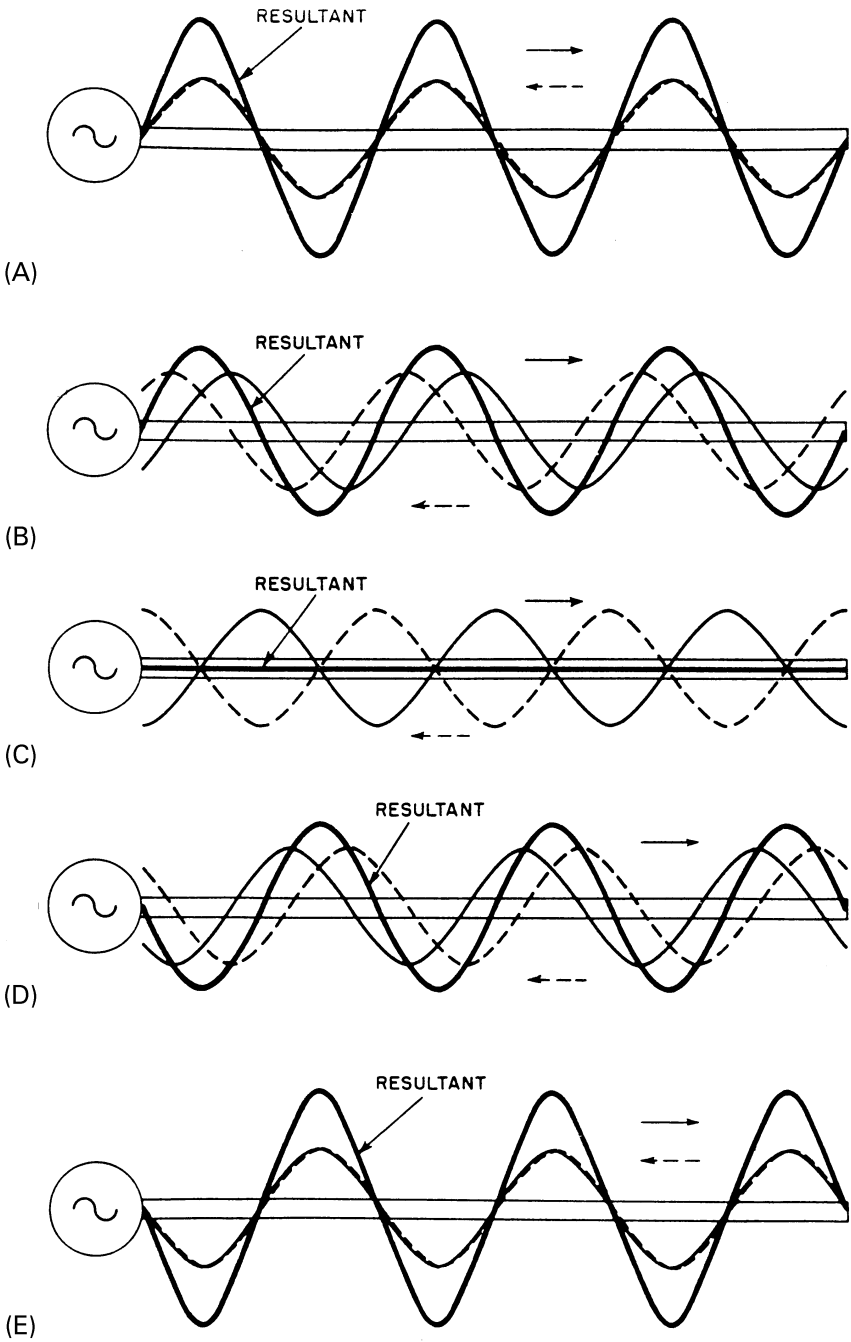
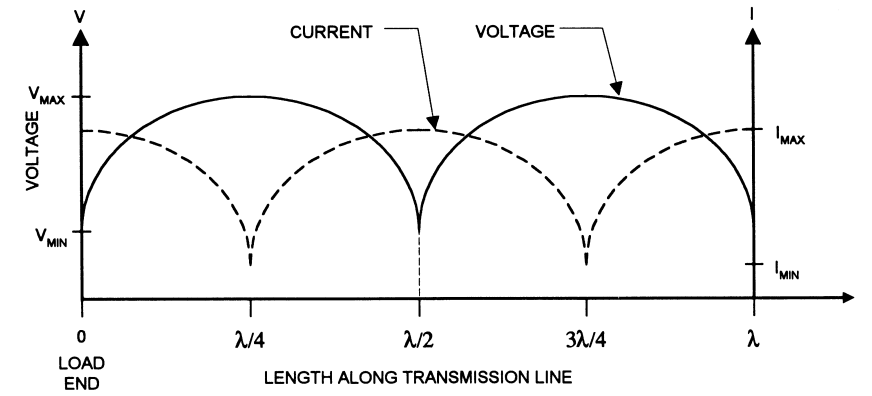
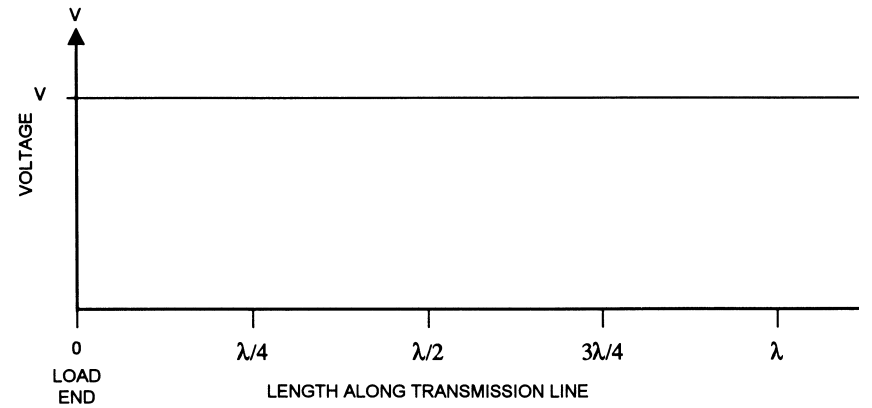

RESULTANT
(A)
RESULTANT
(B)
RESULTANT
(C)
RESULTANT
(D)
RESULTANT
(E)

FIGURE 2.14



(F)



(G)

FIGURE 2.14 *(continued)*

apart in opposite directions in Figure 2.14B, which causes the overall amplitude of the standing wave to decrease, but the location of the maximum and minimum points remain stable. The condition of 180° out of phase between the two traveling waves results in a zero-amplitude standing wave, as shown in Figure 2.14C. No current flows in this case. As the waves continue to move apart, the standing wave reappears, as shown in Figure 2.14D. The final case, Figure 2.14E, is the case with the traveling waves in-phase with each other (0° difference). Notice that the standing wave in Figure 2.14E is the same amplitude as that of Figure 2.14A, but is 180° out of phase with it.

Figure 2.14F shows the voltage along a transmission line when a VSWR higher than 1:1 is present. The antenna or load end is on the left, and the lengths along the transmission line are plotted in terms of wavelength. In the

situation shown, the voltage is a minimum at the antenna end (marked '0'). It rises to a peak or 'antinode' at a quarter wavelength, and then drops back to a minimum ('node') at the half-wavelength ($\lambda/2$) point. The voltage then rises again to a peak at $3\lambda/4$, falling back to minimum at 1λ. The current rises and falls in a similar manner, but the nodes and antinodes are offset from the voltage by 90° (quarter wavelength).

Note that the voltage is at a minimum and the current at a maximum at the same points that are integer multiples of a half wavelength. This situation has the effect of making it advisable to measure the VSWR at the transmitter end only when integer multiples of an electrical half wavelength of transmission line (note: the most valid measurement is made at the interface of the transmission line and antenna feedpoint).

The plot shown in Figure 2.14G is the same line system under the situation where the antenna impedance and transmission line impedance are matched. For obvious reasons, this line is said to be 'flat.'

Measures of SWR

There are several different methods for measuring the SWR of an antenna transmission line system. For example, one could measure the incident and reflected signal voltages along the line, producing a result of $\text{VSWR} = (V_\text{i} + V_\text{r})/(V_\text{i} - V_\text{r})$. The method of Figures 2.14F and 2.14G can be used by comparing the maximum and minimum voltages: $\text{VSWR} = V_\text{MAX}/V_\text{MIN}$. We can also measure the forward and reverse power levels to find the VSWR. A simple way to predict the VSWR is to compare the antenna feedpoint resistive impedance (Z_L) to the transmission line characteristic impedance (Z_0). The value of the VSWR is found from either

$$\text{VSWR} = \frac{Z_0}{Z_\text{L}} \qquad (Z_0 > Z_\text{L})$$

or

$$\text{VSWR} = \frac{Z_\text{L}}{Z_0} \qquad (Z_\text{L} > Z_0)$$

If, for example, we measure the antenna feedpoint impedance as being 25 ohms, and the antenna transmission line is 52 ohm coaxial cable, then the $\text{VSWR} = 52/25 = 2.08{:}1$.

We can find the resonant frequency of an antenna by finding the frequency at which the VSWR is a minimum. But that point may not be a VSWR of 1:1 unless the resistive component of the antenna feedpoint impedance is the same as the transmission line impedance, and the complex portion of the impedance is zero. To create this situation one might need an impedance matching device (see Chapter 12).

AN OLD MYTH REVIVED?

There are a number of myths that are widely held among radio communications hobbyists – and amateur radio is no less infested with some of these myths than others (CB, for example). Twenty-five years ago I worked in a CB shop in Virginia, and we kept hearing one old saw over and over again: you can 'cut your coax to reduce the VSWR to 1' (actually, they meant '1:1' but routinely called it '1'). Hoards of CBers have 'cut the coax' and watched the VSWR reduce to 1:1, so they cannot be talked out of the error. I even know of one shop that kept 30 cm lengths of coaxial cable, with connectors on both ends, so they could insert them into the line at the transmitter in order to find the correct length that would reduce the VSWR to 1:1. What actually happens in that case is a measurement artifact that makes it appear to be true.

Of course, hams are superior to CBers and so do not believe that error, right? I would like to think so; but having been in both the CB and the amateur worlds, and 'Elmered' (mentored) more than a few CBers studying for amateur licenses, I have to admit that at least as many amateurs believe the 'cut the coax' error as CBers (sorry, fellows, but that's my observation).

The only really proper way to reduce the VSWR to 1:1 is to tune the antenna to resonance and then match the impedance. For a center-fed half-wavelength dipole, or a bottom-fed quarter-wavelength vertical, the proper way to resonate the antenna is to adjust its length to the correct point. The formulas in the books and magazines only give approximate lengths – the real length is found from experimentation on the particular antenna after it is installed. Even commercial antennas are adjusted this way. On certain CB mobile antennas, for example, this trick is done by raising (or lowering) the radiator while watching the VSWR meter. On amateur antennas similar tuning procedures are used.

Even when the resonant point is found, the feedpoint impedance may not be a good match to the transmission line. A VSWR will result in that case. The impedance matching should be done between the far end of the transmission line (i.e. away from the receiver or rig) at the feedpoint of the antenna. Antenna tuners intended for strictly coaxial cable are little more than line flatteners. They do not really 'tune' the antenna, but rather they reduce the VSWR looking into the transmission line so that the transmitter will work properly. If the antenna tuner is not a high-pass filter (as some are), then it will also provide some harmonic attenuation.

An approach used by many amateurs (including myself) is to connect an antenna-matching unit (tuner) at the output of the transmitter. For my Kenwood TS-430, I use either a Heath SA-2060A or an MFJ Differential Tuner to 'tune-out' the VSWR presented by my Hustler 4BTV and 23 m of coaxial cable. But I do not even pretend to be tuning the antenna. The

TS-430 is a solid-state rig, and the finals are, therefore, not terribly tolerant of VSWR, and will shut down with a high VSWR. The purpose of the antenna tuner is to reduce the VSWR seen by the transmitter – ignoring the actual antenna mismatch on the roof. The tuner also serves to reduce harmonics further, thereby helping to prevent television interference.

The best form of antenna tuner is one that both reduces the VSWR (for the benefit of the transmitter), and also resonates to the antenna frequency, preventing harmonics from getting out (a little secret is that many 'line flattener' antenna tuning units are actually variable high-pass filters, and must be used with a low-pass filter ahead of them if spurious signals are to be kept in abeyance).

Should you even worry about VSWR on a system? Or, more correctly stated, given that a 1:1 VSWR often requires a herculean effort to achieve, at what point do you declare the battle won and send the troops home? Some of the issues are:

- Transmitter heating of the transmission line due to power losses.
- Reduction of power from solid-state transmitters due to SWR shutdown circuitry.
- At kilowatt levels there may be excessive radio frequency voltage at nodes, which could lead to transmission line shorting.
- Loss of receiver sensitivity. The feedline loss is added directly to the receiver noise figure, so may degrade the matched noise figure of the receiver. This problem is especially severe in the VHF/UHF scanner bands.

Where these problems result in unacceptable performance, work hard to match the antenna to the feedline, and the feedline to the transmitter or receiver. As one of my correspondents said: 'The question then is to what extent does the problem impair the actually ability to use the receiver or transmitter successfully?' The same reader provided the guidelines below:

(1) The actual SWR on a feedline is 1:1 only if the load (e.g. an antenna) connected to the line is equal to the characteristic impedance of the line. Adjusting an antenna to resonance will improve the SWR only if the impedance of the antenna at resonance is closer to the impedance of the feedline than before it was adjusted. If the impedance of the antenna at resonance is not equal to the impedance of the feedline, you will never get the SWR down to 1:1.

(2) The only possible places to measure the SWR with consistent accuracy are:

(a) at the load

(b) at a distance of 0.5 electrical wavelength (accounting for the velocity factor of the line), or integer multiples thereof, from the load.

The latter is true because a half-wavelength long piece of feedline 'repeats' the impedance of whatever is connected to its far end; it is almost as if you are measuring right up at the feedpoint. As a consequence, many professional antenna installers go to pains to make the transmission line $\lambda/2$ or an integer multiple of $\lambda/2$. The VSWR and impedance looking into the line will reflect the situation at the feedpoint. Be aware that the velocity factor of the transmission line reduces the physical length required to make an electrical half wavelength.

(3) An easy way to check if your SWR measurements are being accurately made is as follows: Add a piece ($\lambda/8$ to $\lambda/4$ wavelength) of *identical* line to your feedline. Repeat the SWR measurement. If it is not the same, your SWR measurement is not accurate, i.e. you are not measuring the actual SWR.

(4) Know that a few things can upset SWR measurements, e.g. currents flowing on the outside of a coaxial feedline (a common situation). This can be caused by not using a BALUN when feeding a balanced antenna (e.g. a symmetric dipole) with an unbalanced (coaxial) feedline. It can also be induced by antenna currents onto a nearby coaxial feedline.

ANTENNA GAIN, DIRECTIVITY, AND RADIATION PATTERNS

Antennas do not radiate or receive uniformingly in all directions (and, indeed, you usually do not want them to!). The interrelated concepts of **antenna gain** and **directivity** are of much concern to antenna builders. Gain refers to the fact that certain antennas cause a signal to seem to have more power than it actually can claim. The gain of the antenna can be expressed as either a multiple (e.g. a twofold increase in apparent power) or in decibel notation (e.g. a 3 dB increase). Both methods of expressing gain reflect the fact that a gain antenna seems to produce a signal that is stronger than the signal produced by some comparison antenna (e.g. a dipole). The apparent power produced by the antenna and transmitter combination is called the **effective radiated power** (ERP), and is the product of the transmitter power and the antenna gain. For example, if the antenna gain is 7 dB, the power is increased by a factor of five times. The ERP of a 100 W transmitter is therefore $5 \times 100\,\text{W} = 500\,\text{W}$.

On a receive-only antenna, the signal picked up by a gain antenna is louder than the same signal picked up on a non-gain antenna or the comparison antenna. Because of reciprocity, the gain is the same for both receive and transmit.

Directivity refers to the fact that the radiation or reception direction is not constant. Certain directions are favored and others rejected. Indeed, this is how the antenna gets gain. After all, if there is an apparent increase in the

signal power, there must be some phenomenon to account for it – after all, antennas do not produce power. Figure 2.15A shows how the gain and directivity are formed. Consider an **isotropic radiator** source located at point 'A' in Figure 2.15A. An isotropic radiator is a perfectly spherical

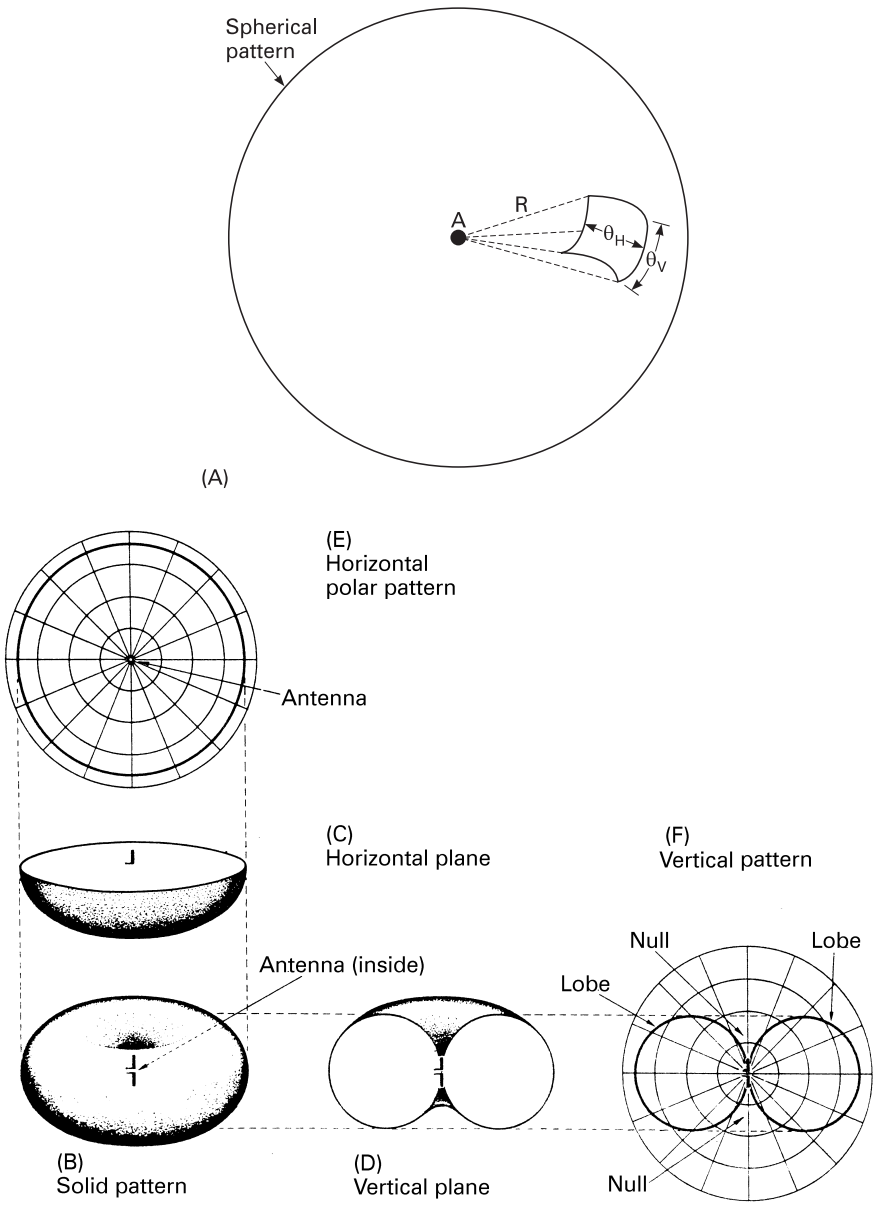


FIGURE 2.15

point source. The signal radiates outward from it in a spherical pattern. At some distance R we can examine the surface of the sphere. The entire output power of the transmitter is uniformly distributed over the entire surface of the sphere. But if the antenna is directive, then all of the power is focussed onto a small area of the sphere with a vertical angle of θ_V and a horizontal area of θ_H. The same amount of energy exists, but now it is focussed onto a smaller area, which makes it appear as if the actual power level was a lot higher in that direction.

The **antenna pattern** is a graphical way of showing the relative radiation in different directions (i.e. directivity). The most commonly seen pattern is the **azimuthal** or **polar** plot. That pattern is measured in the horizontal plane at all angles around the antenna, and is plotted graphically as if seen from above. It is a 'bird's eye view' of the antenna radiation characteristic. But antenna radiation patterns are actually three-dimensional, and the azimuthal plot is merely a slice of that three-dimensional pattern, as seen from but one perspective.

Figures 2.15B through 2.15F show how antenna patterns are developed. The antenna in this case is a vertically polarized Hertzian radiator such as a half-wavelength dipole. It radiates poorly off the ends, and very well at angles perpendicular to the wire. The solid-pattern of Figure 2.15B shows the three-dimensional view of what the radiation looks like. To find the azimuthal and elevation patterns for this antenna, we take slices out of the solid (Figures 2.15C and 2.15D), and plot the results (Figures 2.15E and 2.15F). Because the radiator is vertical (and perfect, I might add), it radiates equally well in all horizontal directions (see Figure 2.15E), so the azimuthal or horizontal plot is termed **omnidirectional**. The elevation or vertical plane view (Figure 2.15F) has to take into account the minima found off the ends of the radiator, so plots as a figure '8.'

If the vertical radiator of Figures 2.15B–2.15F is rotated 90° to form a horizontal dipole, then the radiation patterns remain the same, but are reversed. The omnidirectional pattern of Figure 2.15E becomes the elevation or vertical extent, while the figure '8' pattern of Figure 2.15F is the horizontal or azimuthal extent.

The patterns of Figures 2.15B–2.15F make a rather bold and unwarranted assumption: the antenna is perfect and is installed somewhere in free space. Since most antennas are not in the zone midway between the Earth and Mars, one has to account for the effects of being close to the Earth's surface. Figure 2.16 shows the approximate patterns for horizontally polarized half-wavelength dipoles close to the ground. In Figure 2.16 we see the azimuthal pattern. It is the figure '8,' but is pinched a bit. Notice that the maxima are perpendicular to the wire, and the minima (or 'nulls') are off the ends. One use for this type of pattern is to position the antenna so that it

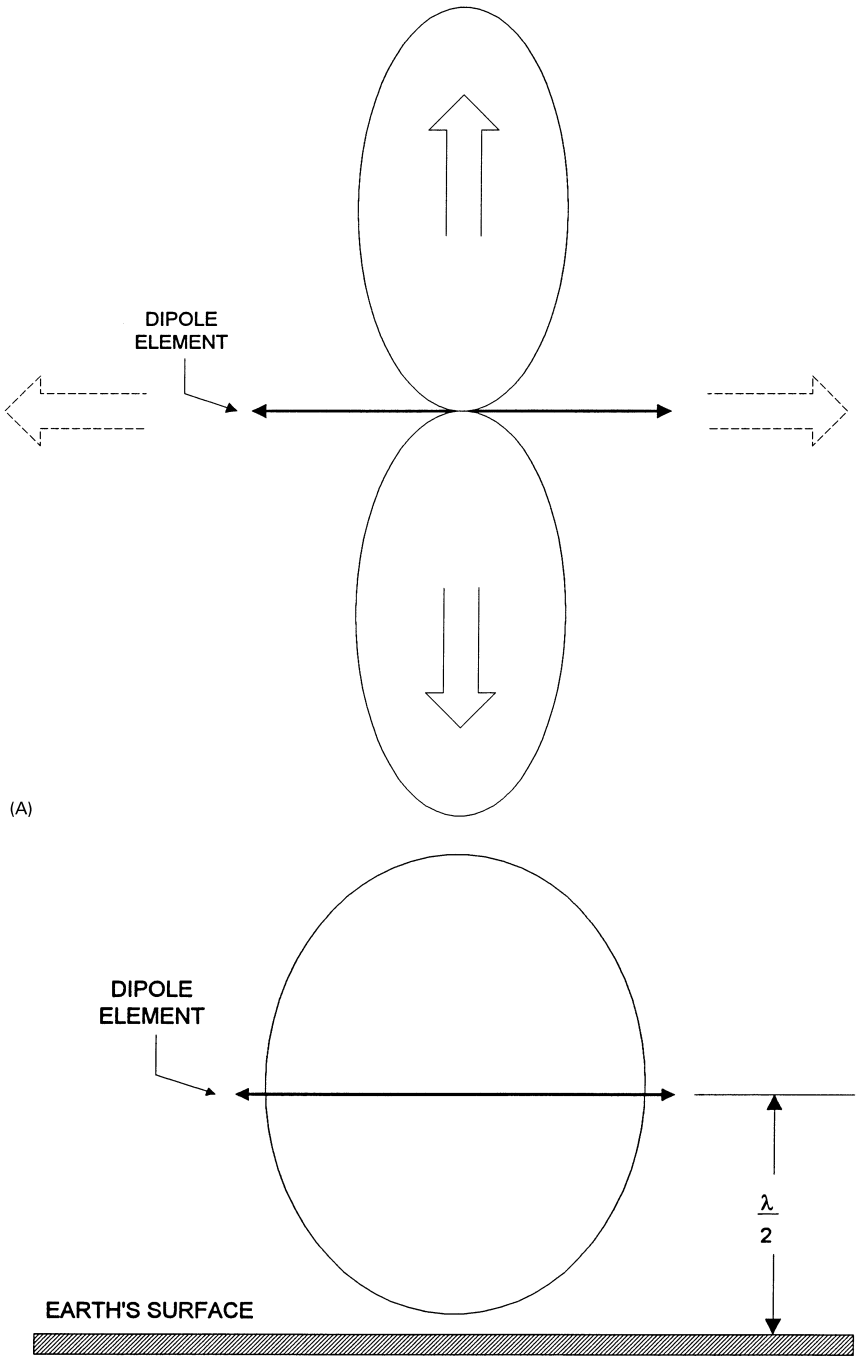


FIGURE 2.16

nulls out interfering signal or noise sources, making the overall signal-to-noise ratio higher – and increasing listenability.

Figure 2.16B shows the elevation extent. This view is taken from a site perpendicular to the wire, and is essentially looking into one end of the three-dimensional figure '8' pattern. In other words, the directivity, or maxima, is in and out of the page. In free space, one would expect this pattern to be omnidirectional. But close to the Earth's surface, reflections from the surface add to the pattern, causing the pinching effect apparent in Figure 2.16B. The shape shown is approximately that for a half wavelength above the ground. Other heights produce variants of this pattern.

The patterns shown in Figure 2.17 are for a vertically polarized Marconi radiator, such as a quarter-wavelength, bottom-fed vertical. The azimuthal pattern, as seen from above, is shown in Figure 2.17A. It is the omnidirectional pattern that one would expect. The elevation pattern in Figure 2.17B

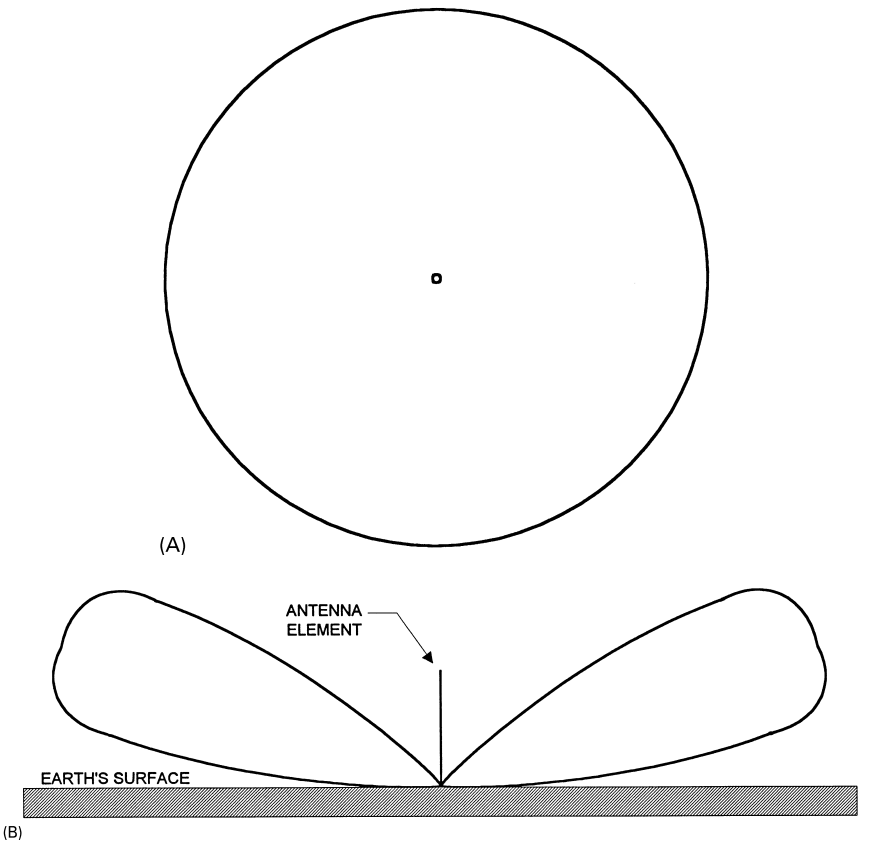


FIGURE 2.17

represents a vertical slice taken from the 'doughnut' seen earlier. Note that the lobes are elevated, rather than being exactly on the horizon. The angle of these lobes is the **angle of radiation**, and affects the distance that an antenna will transmit to or receive from. A low angle of radiation is usually prescribed for long-distance 'DX,' while a high angle of radiation produces skip to nearer regions.

Figure 2.18 shows the conventions for specifying antenna patterns. The antenna is a dipole, with both relative electric field strength and relative powers plotted (both types of plot are done). The 0–270° line is along the length of the dipole, while the 90–180° line is perpendicular to the dipole. The **beamwidth** of the antenna, which is specified in degrees, is measured as the angle between the **half-power points**, i.e. the points where the voltage field drops off to 70.7% of the maximum (along the 90° line), or to the 0.5 power point. These points are also called the '−3 dB' points. Recall that −3 dB represents a power ratio of 1:2.

SOME ANTENNA CONSTRUCTION HINTS

This section will help you think up some good ideas on how to erect the antenna. First, though, let us talk a little bit about antenna safety. It is difficult to say too much about this topic. Indeed, I mention it elsewhere in this book as well as here. Wherever it seems prudent, I warn readers that

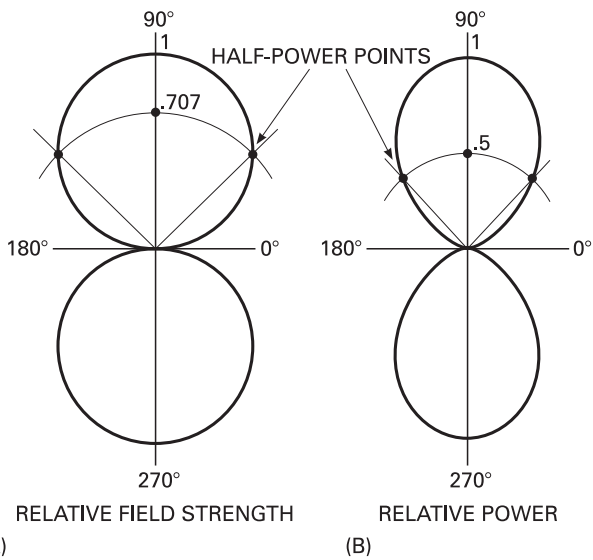


FIGURE 2.18

antenna erection can be dangerous if you do not follow certain precautions. Do it right, and the risk is mitigated and the job can be quite safe.

Safety first!

Before dealing with the radio and performance issues, let us first deal with safety matters – you do not want to be hurt either during installation, or during the next wind storm. Two problems present themselves: reliable mechanical installation and electrical safety.

Electrical safety note. Every year we read sad news in the magazines of a colleague being electrocuted while installing, or working on, an antenna. In all of these tragic cases, the antenna somehow came into contact with the electrical power lines. Keep in mind one dictum and make it an absolute:

There is never a time or situation when it is safe to let an antenna contact the electrical power lines! None. Ever. BELIEVE THIS NO MATTER WHAT ANYONE SAYS TO THE CONTRARY.

This advice includes dipoles and long wires 'thrown over' supposedly insulated power mains lines, as well as antennas built from aluminum tubing. The excuse that the lines are insulated is nonsense. Old insulation crumbles on contact with even a thin wire antenna. Do not do it! The operant word is **never**!

Consider a typical scenario involving a four-band trap vertical antenna made of tubing. It will be 5–8 m tall (judging from adverts in magazines), and will be mounted on a roof, or mast, 4–10 m off the ground. At my home in Virginia, a 7.6 m tall trap vertical is installed atop a 4.6 m telescoping television antenna mast. The total height above ground is the sum of the two heights: 7.6 m + 4.6 m = 12.2 m. The tip of the vertical is 12.2 m above the ground. I had to select a location, on the side of my house, at which a 12.2 m aluminum pole could fall safely. Although that requirement limited the selection of locations for the antenna, neither my father-in-law (who helped install the thing) nor myself was injured during the work session. Neither will a wind storm cause a shorted or downed power line if that antenna falls over.

In some jurisdictions, there might be legal limitations on antenna location. For example, some local governments in the USA have a requirement that the antenna be able to fall over and land entirely on your own property. Before installing the antenna, check local building codes.

When installing a vertical, especially one that is not ground mounted, make sure that you have help. It takes at least two people to safely install the standard HF vertical antenna, and more may be needed for especially large models. If you are alone, then go and find some friends to lend an arm or two. Wrenched backs, smashed antenna (and house parts), and other

calamities simply do not happen as often to a well-organized work party that has a sufficient number to do the job safely.

Mechanical integrity

The second issue in installing antennas is old-fashioned mechanical integrity. Two problems are seen. First, you must comply with local building regulations and inspections. Even though the courts in the USA seem to forbid local governments from prohibiting amateur radio activity (on grounds that it is Federal prerogative), local governments in the USA and eslewhere have a justifiable interest, and absolute right, to impose reasonable engineering standards on the mechanical installation of radio antennas. The second issue is that it is in your own best interest to make the installation as good as possible. View local regulations as the *minimum acceptable standard*, not the maximum; go one better. In other words, build the antenna installation like a brick outhouse.

Both of these mechanical integrity issues become extremely important if a problem develops. For example, suppose a wind or snow storm wrecks the antenna, plus a part of your house. The insurance company will not pay out (in most cases) if your local government requires inspections and you failed to get them done. Make sure the mechanical and/or electrical inspector (as required by US law) leaves a certificate or receipt proving that the final inspection was done. It could come in handy when disputing with the insurance company over damage.

A quality installation starts with the selection of good hardware for the installation. Any radio/television parts distributor who sells television antenna hardware will have what you need. I used Radio Shack stand-off brackets, ground pin, and a 5.8 m telescoping mast. Wherever you buy, select the best-quality, strongest material that you can find. Opt for steel masts and brackets over aluminum, no matter what the salesperson behind the counter tells you. Keep in mind that, although salespeople can be knowledgeable and helpful, you, not they, are responsible for the integrity of the installation. In my own case, I found that the 5.8 m mast was considerably sturdier at 4.6 m than when fully up, so I opted to use less than the full length because the installation is unguyed. Because I have never trusted the little cotter pin method of securing the mast at the slip-up height, I drilled a single hole through both bottom and slip-up segments (which telescope together), and secured the antenna mast with a 8 mm stainless steel bolt. The bolt was 'double nutted' in order to ensure that it did not come loose over time.

The television mast is set on a ground mounting pin/plate that is set into a 76 cm deep (local frost line regulations required only 71 cm) fencepost hole filled with concrete. The top end of the mast was secured to the roof over-

hang of the house (see Figure 2.23). That overhang was beefed up with 5×20 cm kiln-dried lumber that was bolted between two, 60 cm center, roof rafters. I felt it necessary to do that because the roof is only plywood, and the gutter guard is only 2.5×20 cm lumber (and is old), and the soffits are aluminum. There was not enough strength to support a 12 m lever arm, whipping around in a 35 knot wind.

Wind can be a terrible force, especially when acting on the 'sail area' of the antenna through a 7.5–12 m lever arm. A shabby installation will tear apart in wind, causing the antenna to be damaged, damage to the house, and destruction of the installation. That is why I recommend 'brick outhouse' construction methods. Over the 33 years I have been in amateur radio, I have seen a lot of verticals toppled over. Except for a few shabby models that were so poorly built that they should not have been on the market in the first place, all of these failed installations were caused by either poor installation design or poor-quality materials.

Antenna erection methods

In this section we will look at some methods for erecting wire antennas. It is recommended that you also consult Chapter 3 to find out how to make connections to wire antennas and the usual fittings (end insulators, center insulators, BALUN transformers, etc.). Chapter 3 also deals with the type of wire used for antenna construction. The ideas in this chapter are not the be-all-end-all discussion, but rather points of departure of representative methods. Your own innate intelligence can figure out other methods suitable for your own situation (keeping safety foremost in your plans).

Figure 2.19 shows one of the most basic methods for installing a wire antenna. A mast is installed on the roof of the house, or on the side of the building close to the roof. Alternatively, the mast could be replaced by some

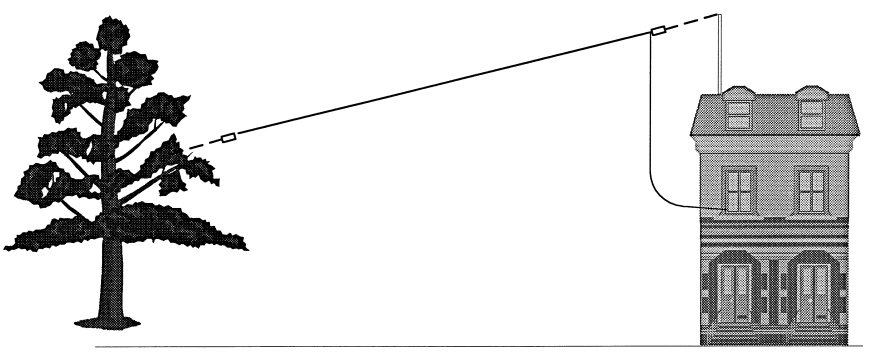


FIGURE 2.19

other support on the house. The other support is a convenient nearby tree. It could also be another building or a mast erected on the ground. The antenna is supported between two end insulators, that are in turn supported by ropes connected to the mast and the tree. For a simple random length Marconi antenna, as shown here, the direction of the wire can be up, down, or horizontal, depending on the convenience of the situation (horizontal is preferred but not always easily attained). The down-lead to the receiver or antenna tuner is run through a convenient window. There are special straps and fittings for making the connection under the window (see your radio dealer).

The installation shown in Figure 2.20 is a half-wavelength dipole, although it could represent a variety of doublet antennas. As with the previous case, the supports can be masts, a building, or a convenient tree. The installation is very much like that in Figure 2.19. The goal in making this installation is to make the antenna horizontal, with the feed-line coming away from the antenna center at a right angle for as far as is practical.

The installation in Figure 2.21 is for a dipole, but it must be a small dipole (either inductively loaded or for a high frequency), or built over a very wide roof. Both end supports are on the roof, or on the walls at either end of the roof. The coaxial cable comes away at a right angle, and then passes through a window. A vertically polarized dipole is shown in Figure 2.22.

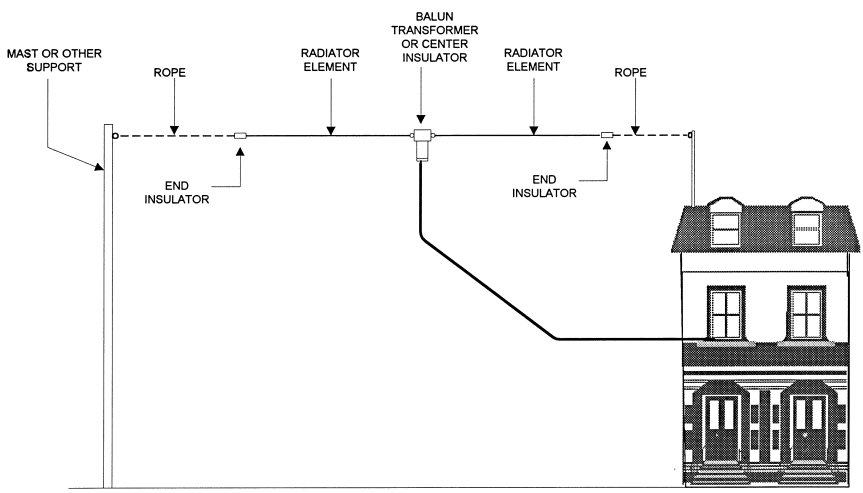


FIGURE 2.20

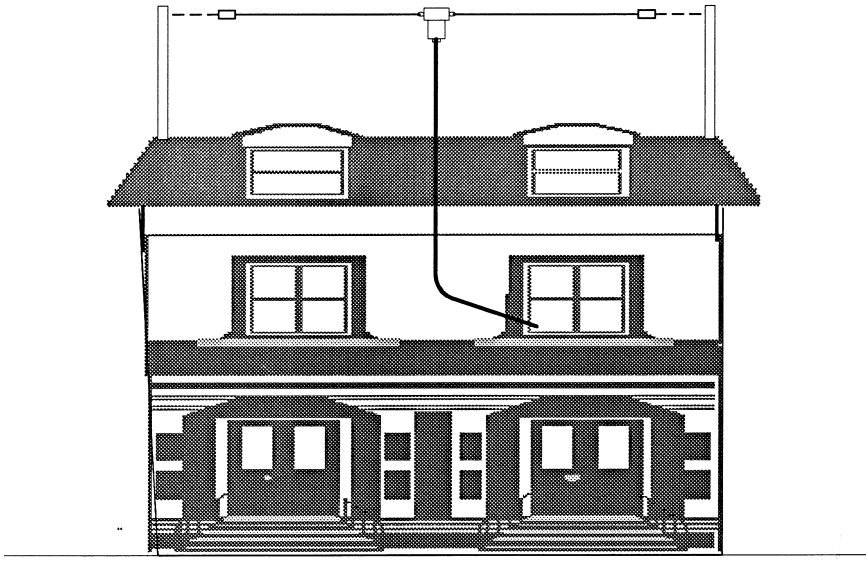


FIGURE 2.21

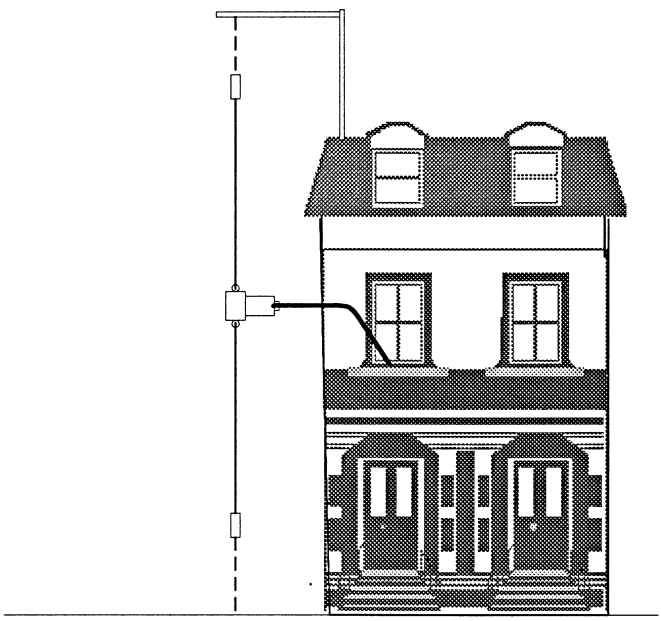


FIGURE 2.22

Figures 2.23A and 2.23B show methods for installing quad loops and delta loops, respectively. As above, the end supports can be a tree, a mast, or a building. It is necessary to support both the top and bottom of loops. The weight of the BALUN transformer and coaxial cable is not sufficient to keep the antenna from whipping around in the wind, and possibly doing some damage to either itself or other structures (not to mention the odd passerby). The bottom support can be either separate ropes to the end supports (Figure

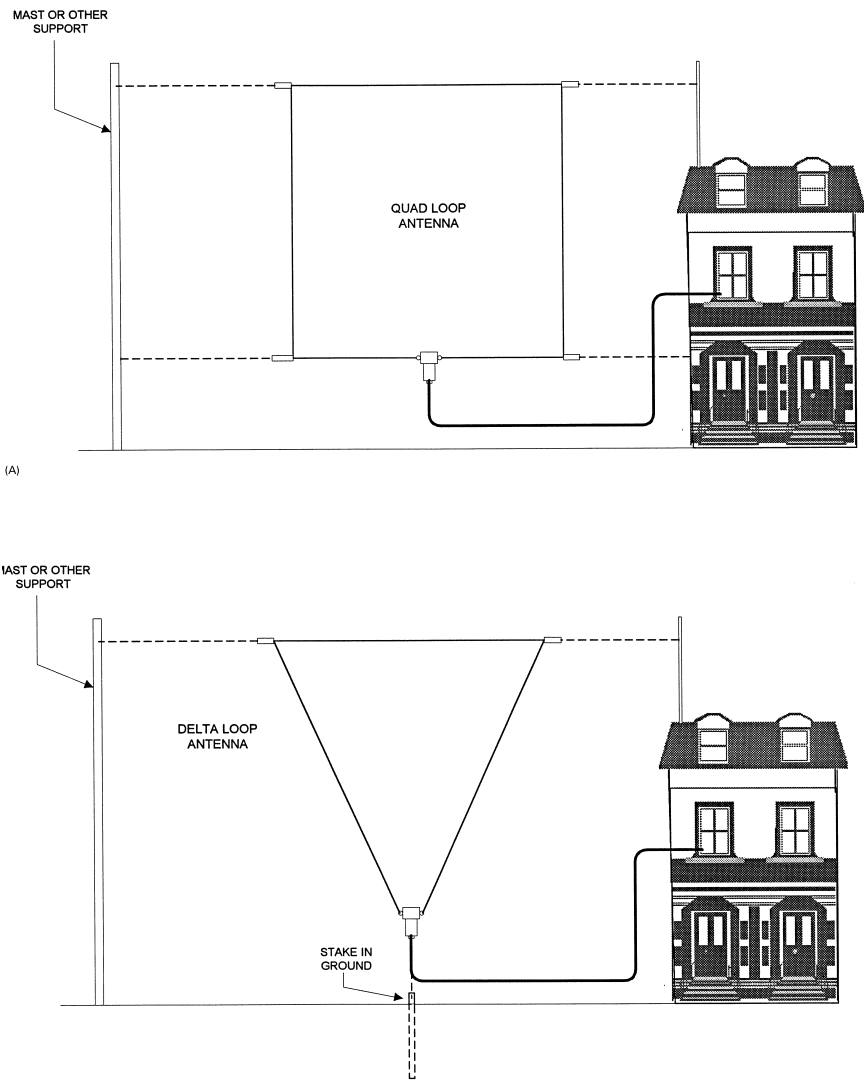


FIGURE 2.23

2.23A), or to a stake in the ground (Figure 2.23B). Make sure that the stake is secure against pulling out under the stress of wind blowing against the sail area of the antenna.

Now, let us go to Chapter 3 to talk about wire and connections.

CHAPTER 3

Wire, connection, grounds, and all that

In this chapter we will take a look at the types of wire useful for antenna elements, the different types of transmission lines, and grounding. The wire used in an antenna project can make or break the results, especially the reliability of the project over time. Antennas, especially outdoor antennas, are exposed to a lot of stress and environmental factors that tend to deteriorate both the performance and life expectancy.

TYPES OF WIRE

There are basically two types of wire used for the radiating element and other parts of antennas: **solid** (Figure 3.1A) and **stranded** (Figure 3.1B). The solid type of wire is made of a single thick strand of copper or copper alloy. Pure copper is rarely used because of both cost and strength problems, while aluminum is almost never used because it does not take solder without special, high-cost, industrial equipment. Solid wire is extruded from a single piece of copper. This is the type of wire used most often in household electrical wiring. In its smaller sizes, it is also the type of wire most commonly used as hook-up wire in electronic construction projects.

The stranded form of wire is made of two or more (sometimes many more) lengths of solid wire twisted together (Figure 3.1B). The number of strands used to make the wire sets the flexibility of the wire. Coarse stranded wire has fewer strands, and is less flexible than wire of the same overall diameter that is made up of many more strands of finer wire.

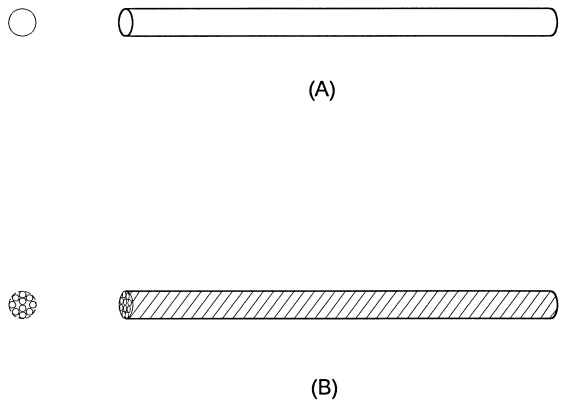


FIGURE 3.1

Antennas should normally be built with stranded wire. The reason is that the stranded wire stands up better to the constant flexing an antenna experiences, especially in windy areas. All wire will begin to stretch because of both gravity and wind. Solid wire will stretch rather rapidly, and eventually come to a point where it is too thin at the strain point to hold its own weight – and will come tumbling down. Stranded wire, on the other hand, withstands the normal stresses and strains of antenna installations much better, and will stay up a lot longer. Stranded wire antennas may come tumbling down as well, but in general it takes a lot more time and effort for the stranded wire antenna to break, compared to the solid wire type.

WIRE SIZES

Both stranded and solid wire comes in certain standard sizes. The specifications for each size wire were set basically for their ability to carry certain standard loads of direct current or the 50/60 Hz alternating current (AC) used in power systems. Table 3.1 gives some of the standard wire sizes in the range normally used by antenna constructors. In most cases, it is a good idea to stick to sizes around AWG 14 (SWG 16) for antenna work. The larger sizes are expensive, and difficult to work (most antenna hardware assumes AWG 14 or SWG 16, unless otherwise marked). The smaller sizes, on the other hand, are nice to use but often provide poor reliability – they break easily. The trade off is to use wires, of whatever standard size, that are in the 1–2.2 mm diameter region.

TABLE 3.1

AWG (USA)	Diameter		Nearest SWG (UK)
	mil[a]	mm	
8	128.5	3.264	10
10	101.9	2.588	12
12	80.8	2.053	14
14	64.1	1.628	16
16	50.8	1.291	18
18	40.3	1.024	19
20	32.0	0.812	21
22	25.3	0.644	24

[a]1 mil $= \frac{1}{1000}$ inch.

COPPER-CLAD WIRE

Pure copper wire is not terribly good for radio antenna work because it is too soft to withstand the rigors of being outdoors, or supporting its own weight plus the weight of the fixtures and transmission lines used. As a result, one needs to find a better wire. Some people will use copper wires of harder alloys, but even that type of wire is not all that good. Others buy 'hard drawn' copper wire, and while it is better than straight electrical wiring copper wire it is not the best suited to antenna construction. The best wire for radio antennas is **copper-clad steel wire**, perhaps the most famous brand of which is Copperweld. This type of wire is made of a solid steel core coated with a copper layer. It looks like ordinary copper wire on the outside. The steel provides strength and thereby increases the reliability of the antenna installation. But steel is not a terribly good electrical conductor, so the manufacturer places a layer of copper over the outside of the wire.

The standard wire sold with the better antenna kits and by the better sources is copper-clad steel, stranded wire of AWG 14 (SWG 16) size. Each strand is copper-clad steel, and the strands are wound together to form the antenna wire.

The copper-clad wire is better for the purposes of strength, but there is a workability problem with it: copper-clad steel-core wire knots easily. If you get a knot or kink (even a minor bend) in it, that feature is always present from then on. Even if you attempt to straighten it out, it will not come anywhere near the original state. Although ordinary copper wire also possesses this attribute, it is not nearly as severe as with copper-clad steel-core wire, and can often be worked into a negligible size when it does occur. The trick is to make sure that the wire does not knot or kink during the installation process (and that can be a monumental task!).

One fair question regarding copper-clad steel-core wire concerns its resistivity. After all, any basic electrical text tells us that the cross-sectional area of the wire sets the current-carrying capacity. That is true for direct current (DC), and power line frequency AC, but only partially true for higher-frequency AC. Radio signals cause AC currents to be set up in the antenna wire, and because their frequency is so much higher than power line AC the currents flow only near the surface. The operating principle here is called the **skin effect**.

SKIN EFFECT

The skin effect refers to the fact that AC tends to flow on the surface of a conductor. While DC flows in the entire cross-section of the conductor, AC flows in a narrow band near the surface (Figure 3.2). Current density falls off exponentially from the surface of the conductor toward the center (inset to Figure 3.2). At the **critical depth** (δ), also called the **depth of penetration**, the current density is $1/\varepsilon$, or $1/2.718 = 0.368$, of the surface current density. The value of δ is a function of operating frequency, the permeability (μ) of the conductor, and the conductivity (s).

WIRE SIZE AND ANTENNA LENGTH

The size of the wire used to make an antenna affects its performance. For amateur radio operators, one concern is the current-carrying capacity. For normal amateur radio RF power levels, the AWG 14 wire size works well,

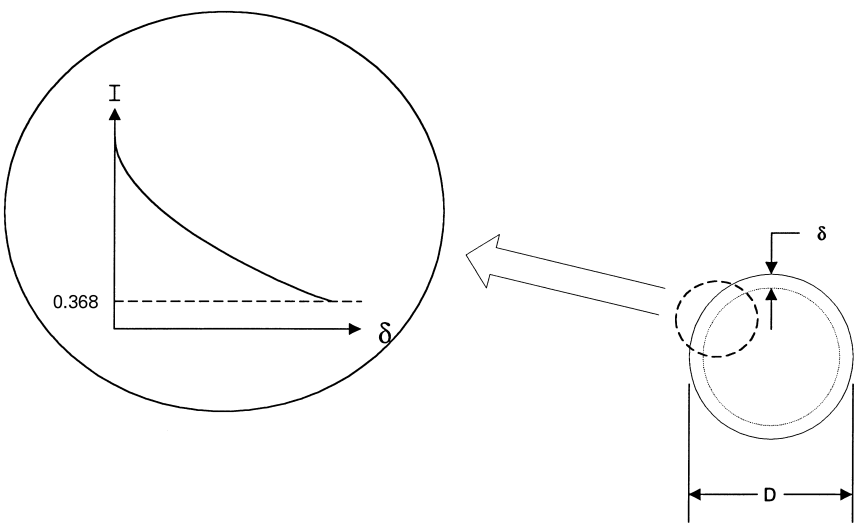


FIGURE 3.2

although those tempted to use smaller diameters with the higher legally permitted powers are forewarned: *it is not such a good idea.*

Both short-wave listeners and amateur radio operators need to be aware of another problem with the wire: the **length to diameter ratio**, or as it is sometimes specified, the **signal-wavelength-to-wire-diameter ratio**, determines the **velocity factor** (VF) of the conductor. The VF is defined as the velocity of the wave in the medium (i.e. wire) compared with the velocity in free space (which is c, the speed of light).

Figure 3.3 shows the velocity factor as a function of the L/D ratio. Wire antennas tend to have very high L/D ratios, especially at high frequencies. For example, a half wavelength at 10 MHz is 15 m (or 15 000 mm), while AWG 14 wire is about 1.63 mm diameter, resulting in $L/D = 15\,000/1.63 = 9200$. The VF of this L/D ratio is around 0.97–0.98. At very high/ultrahigh frequencies (VHF/UHF), on the other hand, the L/D ratio declines. At 150 MHz, the AWG 14 wire half-wavelength antenna has a physical length of 1500 m, making $L/D = 1500/1.63 = 920$. The VF drops to about 0.95–0.96. At 450 MHz, the L/D ratio drops to 200, and the VF runs to 0.94 or so.

So what? Why is the VF important in antenna design? Most of the antennas in this book are designed to be resonant, which means that the element lengths are related to wavelength: quarter wavelength ($\lambda/4$), half

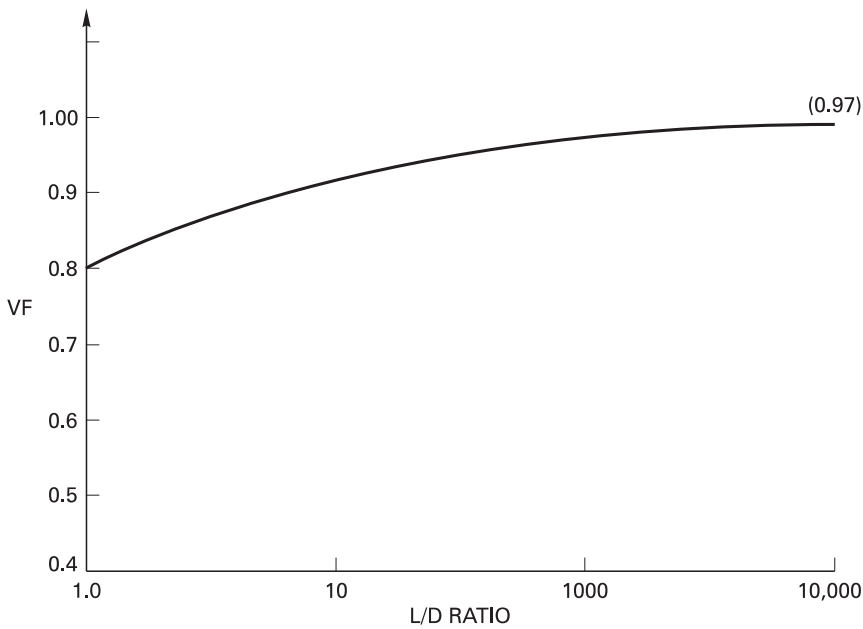


FIGURE 3.3

wavelength ($\lambda/2$), full wavelength (λ), and so forth. These lengths are required electrical lengths, and are only found in free space with perfect connectors. In antennas and transmission lines, the VF shortens the physical length required to achieve a specified electrical length. To find the physical length, multiply the required electrical length by the velocity factor. For example, a half wavelength in free space is defined as $150/F_{\mathrm{MHz}}$, while in a wire it is $(150 \times \mathrm{VF})/F_{\mathrm{MHz}}$. The result is that antenna elements must be shortened a small amount compared to the free space length. If you cut some antennas to a physical length that is comparable to the wavelength of the desired resonant frequency, then you will find that the actual resonant frequency is a tad low – meaning that the antenna is a little bit too long. This effect is almost negligible on the 80 m band, but becomes more substantial at 10 m and downright annoying at 2 m.

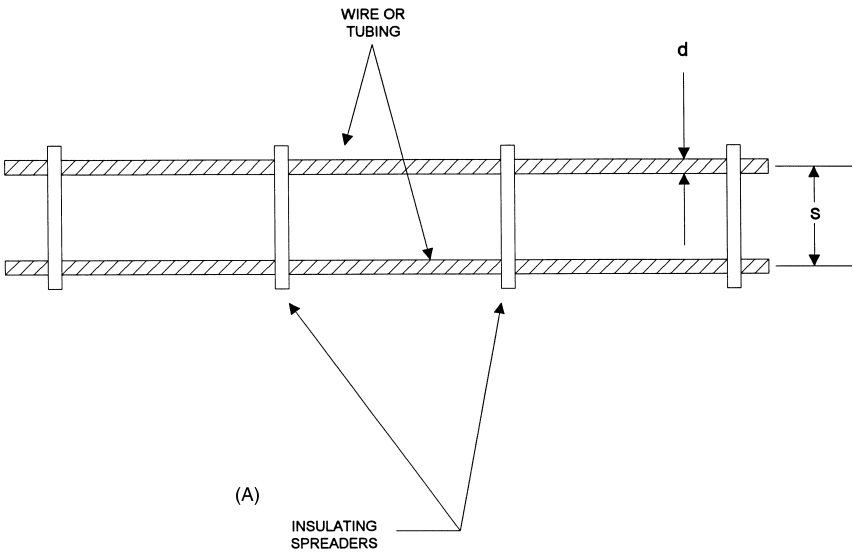
TRANSMISSION LINES

The wire between the antenna and either a transmitter or receiver is called the **transmission line**. Although it is tempting to think of the transmission line as a mere wire, it is actually equivalent to a complex inductor–capacitor network. The details of the transmission line network are a bit beyond the scope of this book, but can be found in antenna engineering textbooks for those who are interested. The main aspect of the transmission line network that you need to understand is that the line possesses a property called the **characteristic impedance** or **surge impedance**, which is symbolized by Z_0. The rigorous definition of Z_0 can get a bit involved, but it breaks down to one thing for most practical situations: Z_0 is the value of load impedance that will result in the maximum power transfer between the antenna and the transmission line, or between the receiver/rig and transmission line. The act of making these impedances equal to each other (a very desirable thing, as you will find out) is called **impedance matching**, and is covered in Chapter 11.

Transmission lines come in a bewildering variety of species, but there are only a few basic types: **parallel open lead**, **twin-lead**, and **coaxial cable** are the main types that we need to consider. Figure 3.4 shows all of these varieties.

Parallel open-wire line

Parallel line is shown in Figure 3.4A. It consists of a pair of wires, run parallel to each other, and separated by a constant space. Insulating spacers (made of ceramic, nylon, or some other material) are used in practical parallel open line to keep the distance between the conductors constant. The characteristic impedance of open line is determined by the diameter (d) of the conductors and the spacing (s) between them. Typical values run from 300 to 1000 ohms. Although you can calculate the spacings needed



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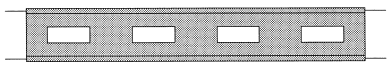

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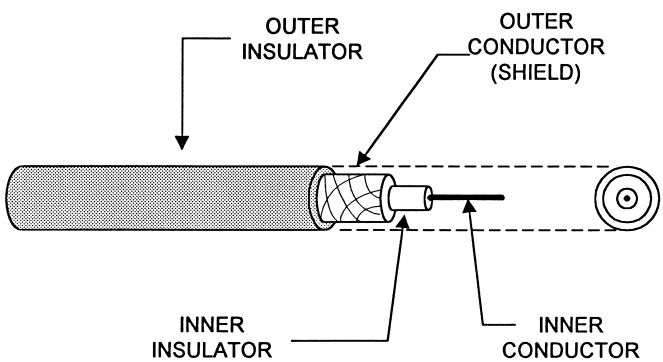


(D)

FIGURE 3.4

(and the wire diameters are given in some of my other books), there are some general guidelines in Table 3.2. The conductor type is along the vertical axis on the left, and the characteristic impedances are along the horizontal axis at the top. The entries in each cell are the spacings (s) in centimeters. As you can see from the table, only some of the calculated values are useful (who wants to make a 129 cm wide parallel line?).

Parallel open-wire transmission line can be either built or bought. For some applications, such as tuned feeders, it might be the best selection. But for many balanced antennas that use parallel line one might want to consider using twin lead instead.

Twin-lead line

The twin-lead form of transmission line is shown in Figures 3.4B and 3.4C. This type of line is like parallel line, except that the conductors are buried in an insulating material that also spans the space between the conductors. Spacings are shortened a little bit compared to the open-wire variety because the dielectric constant of the insulating material reduces the VF.

The most common form is 300 ohm television antenna twin-lead, of the sort shown in Figure 3.4B. This line is about 1 cm wide, and has a characteristic impedance of 300 ohms. Television twin-lead is easily available from a wide variety of sources, including shops catering for television and video customers. It can be used for transmitting at powers up to about 150 W, but even that level is questionable if the line is of a cheaper variety. Receive-only installations have no such limitation.

Television 300 ohm twin-lead may be solid as shown, or may have oblong or rectangular holes cut into the insulation separating the conductors. This tactic is used to reduce the losses inherent in the transmission line. In other cases, the center insulating material will be rounded and hollow. This

TABLE 3.2

	Characteristic impedance, Z_0 (ohms)				
Conductor	300	400	500	600	800
8 AWG	2.0	4.6	10.6	24.4	129
10 AWG	1.6	3.6	8.4	19.3	102
12 AWG	1.25	2.9	6.7	15.3	81
14 AWG	1.0	2.3	5.3	12.2	64
16 AWG	0.80	1.8	4.2	9.6	51
18 AWG	0.63	1.4	3.3	7.6	40
20 AWG	0.50	1.1	2.6	6.1	32
22 AWG	0.40	0.90	2.1	4.8	25

approach to design also reduces the losses, and is especially effective at UHF.

The type of twin-lead shown in Figure 3.4C is 450 ohm line. This type of line is specialist, and must be sought at shops catering for amateur and commercial radio operators. It is about twice as wide as the 300 ohm variety, and almost always comes with rectangular or oblong loss-reducing holes cut into the insulation. The 450 ohm twin-lead is considerably heavier than 300 ohm line, and uses larger conductors. As a result, it can handle higher power levels.

Twin-lead is wonderful stuff, but can cause problems in some installations. The problems are especially severe at VHF/UHF, but nonetheless exist at HF as well. For example, care must be given to where the line is run: it should not be in close proximity to metal structures such as aluminum house siding, rain gutter downspouts, and so forth. It is also susceptible to picking up local electrical fields, which means a possibility of interference problems. One solution is to use coaxial cable transmission line.

Coaxial cable

This type of cable is shown in Figure 3.4D. The name 'coaxial' comes from the fact that the two conductors are cylindrical (the inner one being a wire) and share the same axis. The inner conductor and outer conductor are separated by the inner insulator. The outer conductor is usually a braided wire or foil shield, and is covered with an outer insulating material.

The characteristic impedances of commercial coaxial cables are of the order of 36–120 ohms, with 52 and 75 ohms being the most common. The specific impedance value is a function of the conductor sizes and the spacing between them. There are several standard types of coaxial cable, and these are available in many subtypes (Table 3.3) that use similar numbering.

The terms 'thick' and 'thin' would have been replaced with actual dimensions if this book were being written only a few years ago, but because of the wide array of different types now available it was decided to use these designations instead. One reason is that there are some RG-8/U specialist cables used by amateur radio and commercial antenna installers that are smaller than regular RG-8/U but larger than RG-58/U or RG-59/U.

TABLE 3.3

Type	Characteristic impedance, Z_0 (ohms)	Size
RG-8/U	52	Thick
RG-11/U	75	Thick
RG-58/U	52	Thin
RG-59/U	75	Thin

One of the differences between the different forms of coaxial cable is the insulating material used for the center insulator. Several types are used: **polyethylene**, **polyfoam** and **Teflon**. A principal effect of these materials is to change the VF of the line.

Velocity factor

The VF (denoted by V in some textbooks) of transmission line is calculated in exactly the same way as the VF of other conductors: it is the ratio of the velocity of the signal in the transmission line over the velocity of the signal in free space. The free space velocity is the speed of light (c), or about 300 000 000 m/s. The VFs of several popular transmission lines are shown in Table 3.4.

The effect of the VF is to reduce the physical length of the coaxial cable that is cut for a particular frequency. And this foreshortening can be quite drastic. For example, if polyethylene cable is cut for an electrical half wavelength at, say, 5 MHz, the actual length $L = (150 \times 0.66)/5 = 19.8$ m, while the wavelength of the signal in free space is 1.52 times longer, or 30 m.

CONNECTING THE TRANSMISSION LINE TO THE RIG OR RECEIVER

The method used to connect the transmission line to a receiver or transmitter depends on the type of output/input connector that is provided for the antenna, and the type of transmission line. In the discussion to follow, the illustrations show a receiver, but apply equally well to transmitters. The only difference is that parallel output on transmitters tends to be large ceramic feedthrough insulators, while on receivers, parallel line inputs tend to be small screw terminals on a plastic, Bakelite or other insulating carrier.

Figure 3.5 shows the basic coaxial connector on a receiver. Although an SO-239 'UHF' connector is shown, and may well be the standard, you may also find BNC, Type-N, or RCA 'phono' plug connectors used on some rigs (I've even seen phono plug connectors used on 100 W transmitters, but that

TABLE 3.4

Type of line	VF
Parallel open-wire	0.95–0.99
300 ohm twin lead	0.82
300 ohm twin lead (with holes in insulation)	0.87
450 ohm twin lead	0.87
Coaxial cable	0.66–0.80
Polyethylene	0.66
Polyfoam	0.80
Teflon	0.72

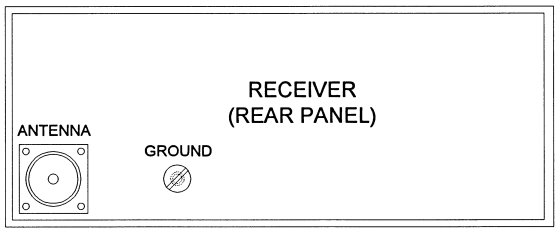
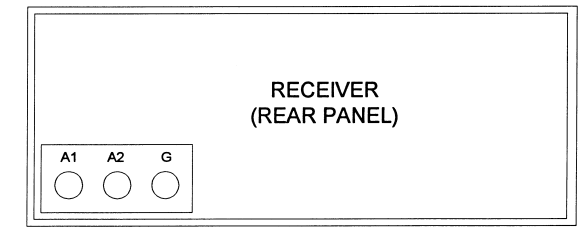


FIGURE 3.5

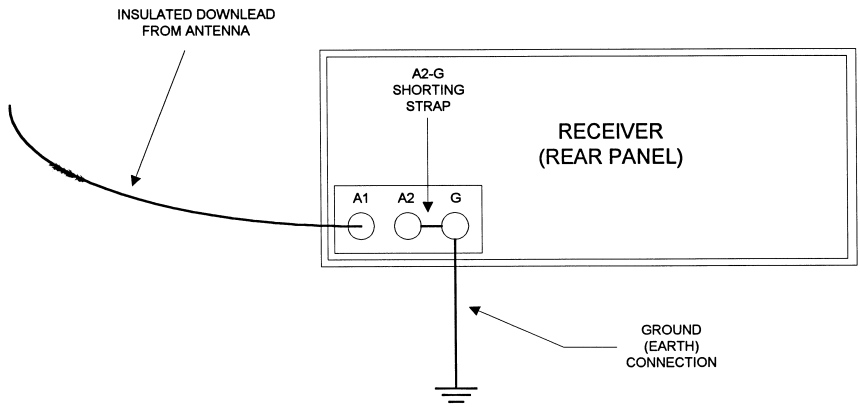
is not the recommended practice!). In this type of installation, the coaxial cable from the antenna (or any antenna tuning unit that might be present) is fitted with a connector that matches the chassis connector on the receiver. The threads of the male PL-259 connector are tightened onto the female SO-239 connector. Or, in the case of BNC connectors, they are pressed on and then rotated a quarter of a turn (it is a bayonet fitting), and phono plugs are just pushed into the socket.

If the receiver or transmitter is fitted with a ground connection, then it should be used. In short order we will discuss what a 'good ground' means in this context, but for the time being rest assured that you will need a heavy wire to the 'good ground' (whatever it is). If your receiver or transmitter lacks a ground terminal, then wonder why and consider that next time you select a new model.

Although most modern receivers are fitted with coaxial connectors for the antennas, there are a few models (including many that are old but still useful) which are fitted with a **balanced input** scheme such as that shown in Figure 3.6. In the case of Figure 3.6A, there are three screw terminals: A1, A2, and G (or the variants 'ANT1', 'ANT2', and 'GND'). The A1 and A2 terminals form a balanced input for a parallel transmission line, while G is the chassis ground. When an unbalanced antenna transmission line, such as a single-wire downlead (Figure 3.6B) is used, then the line goes to either A1 or A2, while the other is connected to G (in the case shown, A1 receives the antenna downlead, while A2 and G are connected together and then connected to the earth ground). Some models are equipped with only two terminals (Figure 3.6C), and in that case the selection is easy: connect the downlead to A1 (or just 'A' or 'ANT'), and the ground lead to 'G' or 'GND.' When parallel line, twin lead, or any other form of balanced line is used (Figure 3.6D), then connect 'G' or 'GND' to ground, and the two conductors of the transmission line to A1 (or ANT1) and A2 (or ANT2).



(A)



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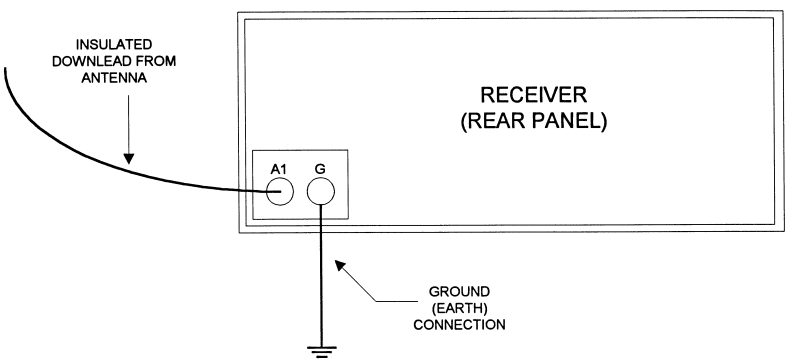

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FIGURE 3.6

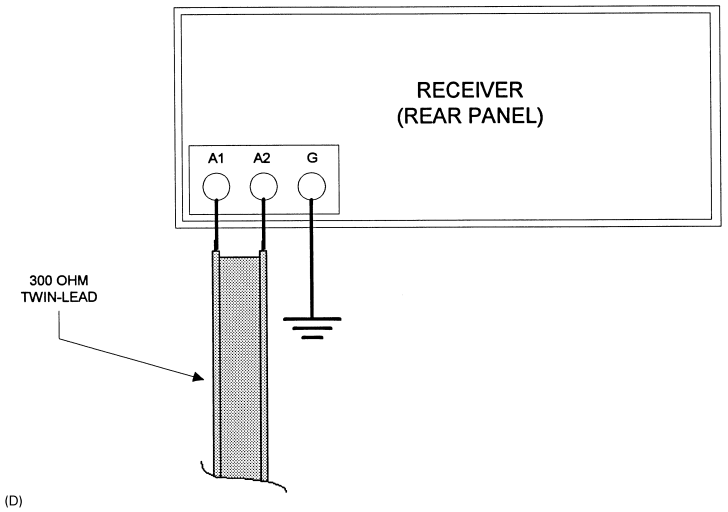


FIGURE 3.6 (*continued*)

CONNECTING WIRES TOGETHER

Wire antennas often require that two or more wires be spliced together (for example, two antenna elements or an antenna element and a downlead). There are right ways and wrong ways to accomplish this task. Let us look at some of the right ways.

All of the correct methods (only a few of which are shown here) have two things in common: they are (a) electrically sound, and (b) mechanically sound. Electrical soundness is created by making a good electrical connection, with the wires bound tightly together, and protected from the elements. The electrical soundness of the joint is made better and longer lasting by soldering the connection.

Figure 3.7 shows how to make the connection mechanically sound. Overlap the two wires being spliced by a few centimeters (Figure 3.7A), and then wrap each one onto the other (Figure 3.7B), forming seven or more tight turns (Figure 3.7C). Once the splice is made, then solder both knots to form a better electrical connection.

One mistake made by novice antenna builders is to assume that the purpose of soldering is to provide mechanical strength. This is false. Solder only improves and keeps good the electrical connection. Solder for radio work is a 50/50 or 60/40 lead/tin mixture with a resin core (NEVER acid core!). It is soft and has no inherent mechanical strength. Always depend on the splice to provide mechanical strength, and NEVER use solder for strength.

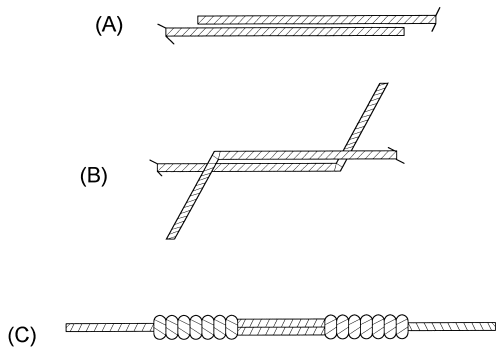


FIGURE 3.7

CONNECTING WIRES TO ANTENNAS AND SUPPORTS

The methods for connecting wires and antenna elements are basically the same as above, but with differences to account for the types of hardware being used in the antenna construction. Figure 3.8 shows how the ends of the wire antenna are supported. **An end insulator** is used. These insulators may be of glass, ceramic, or a synthetic material (nylon is common), but all will have a body and two holes. The wire from the antenna element is passed through one hole, and then looped back onto itself to form seven or more turns. As in the previous case, the splice may be plated with solder to prevent corrosion from interfering with the electrical integrity of the connection. The other end of the insulator is fitted to a rope that is run to a tree, edge of the building, or a mast of some sort. The rope should be treated similarly to the wire in that it should be wrapped over on itself several times before being knotted.

If you use an antenna that has a single-wire downlead, then you should connect it to the antenna in the manner of Figure 3.9. The downlead is almost always insulated wire, even when the antenna is not insulated. The insulation of the downlead prevents it being shorted out somewhere along its run. The scheme in Figure 3.9A shows how most people connect the downlead (and, indeed, how I have connected them quite often). This is not the best practice, although it will work most of the time. The problem is that the connection is stressed by the downlead, and may break. More

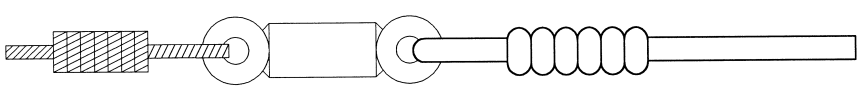


FIGURE 3.8

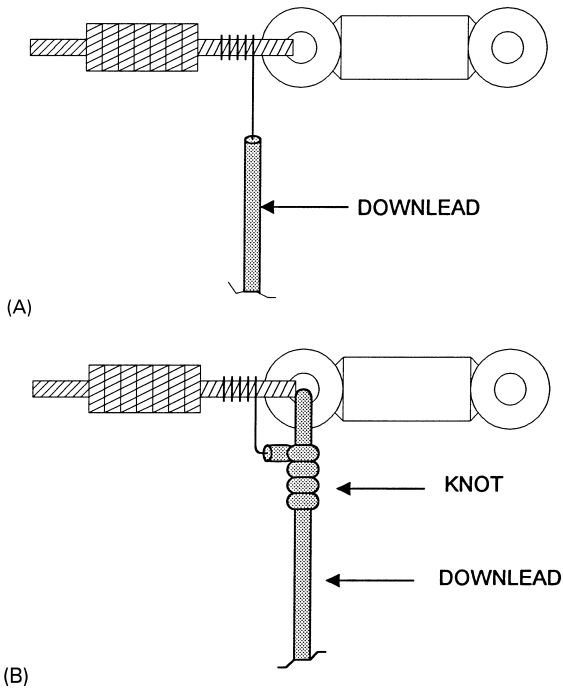


FIGURE 3.9

than once I have been off-the-air due to a spliced downlead breaking right at the junction with the antenna wire. A better solution is shown in Figure 3.9B. In this installation, the downlead wire is passed through the hole in the insulator, and then wrapped back on itself three or four times before being spliced to the antenna wire in the usual manner. This arrangement provides some strain relief to the downlead, and that (usually) translates into a longer life expectancy.

Some antennas, such as the center-fed half-wavelength dipole (a very popular wire antenna), use a center insulator to separate the two halves of the antenna. Coaxial cable is connected such that the inner conductor is spliced to one antenna element, and the shield (outer conductor) is connected to the other antenna element. Good practice is to splice the shield to the element closest to the house, if this is a factor. The method shown in Figure 3.10A will work, but it is also a bit foolish (I know, I know, I have used it and even recommended it in the past). The problem is that the center conductor is quite weak, and may be even smaller than single-wire downleads. As a result, it has no strength. To make matters worse, coaxial cable weighs more length-for-length than single-conductor downleads (or even some twin lead). The connection of Figure 3.10A will work, but only for

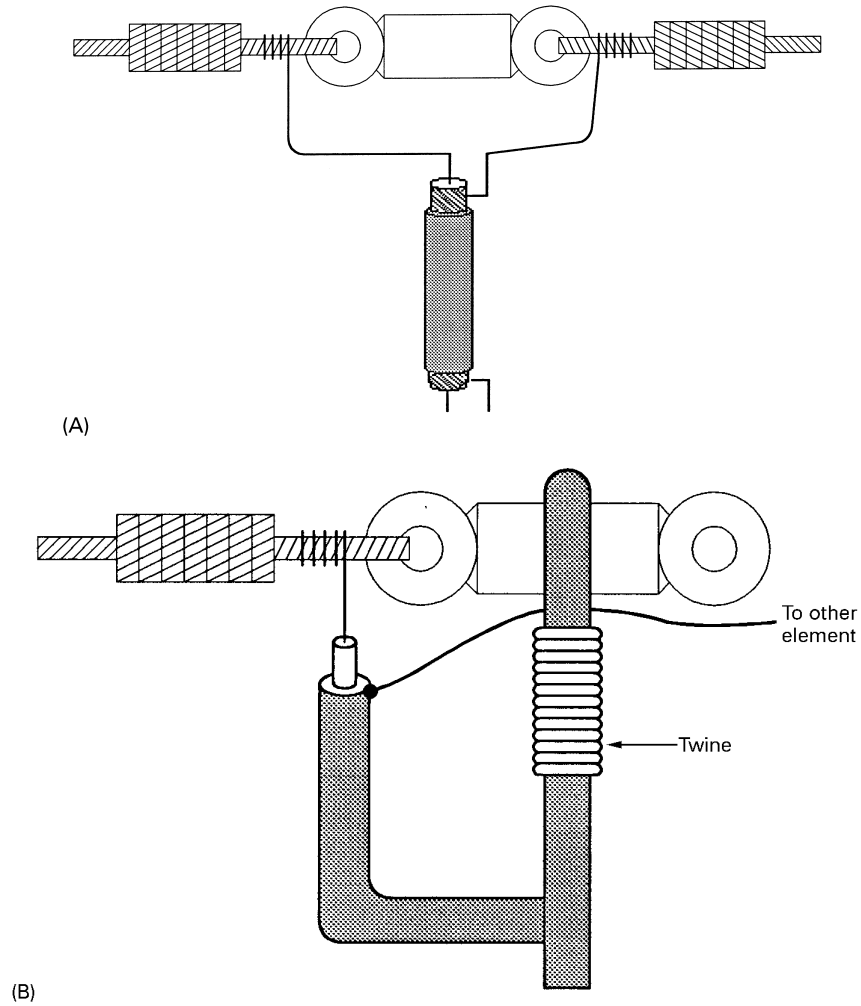


FIGURE 3.10

a while – it will surely come down prematurely. A better scheme is shown in Figure 3.10B. In this case, the coaxial cable is wrapped once around the center insulator, and secured with a length of twine or string. The electrical connections are made in exactly the same manner as shown in Figure 3.10A.

Perhaps the best solution is to use a manufactured center insulator (Figure 3.11). These devices come in a wide variety of sizes and shapes, so this illustration represents a lot of different devices. What they all share in common, however, are means for strain relieving the antenna element wires and coaxial cable. Some devices may be straight insulators, while others may

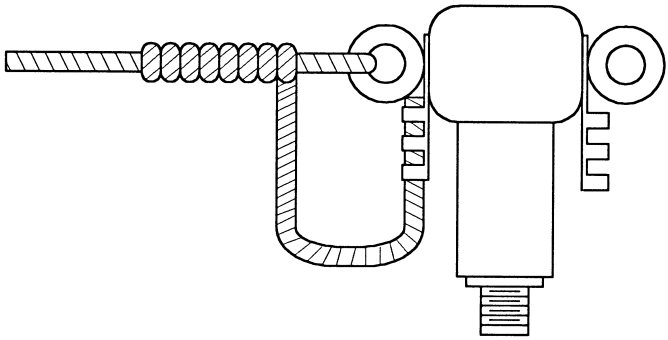


FIGURE 3.11

contain BALUN transformers inside (of which, more shortly). Some models include special strain relief features for the coaxial cable, or they may have a screw-eye on top to secure the antenna center to a mast or other support.

The electrical connections between the center insulator and the wire antenna elements are done in the usual manner (see above), but the ends of the wires are usually attached to eyelet connectors, which are crimped and then soldered. It has been my experience that these connectors are often corroded (as seen by the dull appearance) when they arrive, so brighten them with sandpaper or steel-wool before attempting to make the connections permanent.

LIGHTNING ARRESTORS

Lightning is messy stuff, and can create havoc with your radio – and might set your house on fire. But there are some things you can do to protect yourself. But first, let us take on one myth right away: contrary to popular belief, *antennas do NOT inherently attract lightning*. Lightning will usually strike the highest object around, and if that happens to be an antenna, then the antenna will take a hit. And that hit could set the house on fire, and the radio will be damaged or even destroyed. The lightning would have come anyway, and if it takes out the antenna rather than your house, then it acted like a lightning rod and protected your property.

If lightning actually strikes your antenna, then it will probably wipe out your transmitter or receiver no matter what you do. But a direct hit is not needed for fatal damage to occur: even lightning overhead or striking nearby will induce a large enough voltage into your antenna to wreak havoc inside the equipment. And it may be silent damage because it is not so spectacular as the problems seen with direct hits.

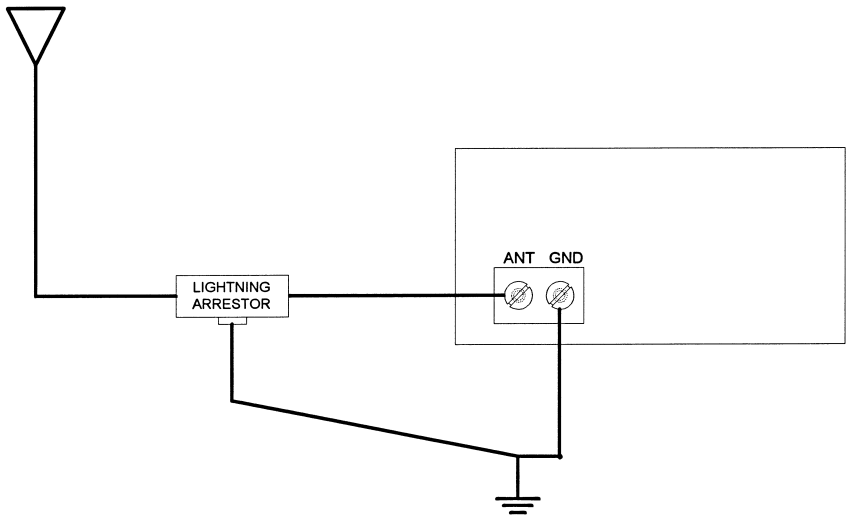
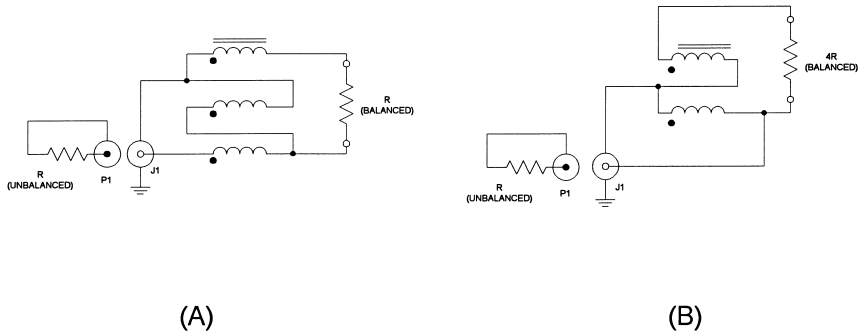


FIGURE 3.12

The key to protecting your equipment is to use a **lightning arrestor** (Figure 3.12). These devices come in several types, each suitable for coaxial line, parallel line, or single-wire downleads. Each will have at least one connection for the antenna wire, and a ground connection. It is placed in the line between the antenna lead and the receiver or transmitter, and should be mounted outside the house. There is very little that can be done for a direct hit, but I have heard tales of near hits only causing a small amount of rig or receiver damage because an effective lightning arrestor was in place. Also, keep in mind that some local laws or insurance regulations make the lightning arrestor mandatory. If you fail to use one, you may be liable for a fine from the local government, or find your homeowner's insurance not valid after a lightning episode. Lightning arrestors are cheap, and so make very reasonably priced insurance.

BALUN TRANSFORMERS

Both impedance transformation and conversion between unbalanced and balanced lines can be accomplished with a BALUN transformer (the name comes from **BALanced–UNbalanced**). Two sorts of transformers are used (Figure 3.13). The 1:1 BALUN transformer converts the balanced load to an unbalanced form (so it can be fed with coaxial cable). The impedance of the load must be the same as the source. The version shown in Figure 3.13B is a 4:1 BALUN. This transformer is used to provide not only a balanced-to-



(A) (B)

FIGURE 3.13

unbalanced conversion, but also a 4:1 impedance transformation. For example, a half-wavelength folded dipole antenna is balanced with respect to ground, and has a feedpoint impedance near 300 ohms. A 4:1 BALUN transformer converts the 300 ohm load to 75 ohms, and makes it appear an unbalanced rather than balanced load.

BALUN transformers can be easily built using toroidal coil forms or ferrite rods. The 1:1 BALUN is wound in the trifilar style, i.e. the three wires are wound side by side so that all three coils interleave with one another. The 4:1 BALUN is wound in the bifilar manner. If you do not wish to make your own BALUN transformers, then you will find that there are many commercially available BALUN transformers on the market.

VHF/UHF scanner users will find few BALUNs specially for their needs, but there are plenty of devices available. The television antenna transformers are not usually called 'BALUNs,' but when you find a transformer that converts 75 ohm coaxial cable to 300 ohm twin-lead, then it is (rest assured) a BALUN, and will work over the entire VHF/UHF range unless marked otherwise. BALUN transformers are bidirectional devices, so can be used either way around.

SAFETY

Antenna safety is an absolute must. First, let me make one thing absolutely, positively, clear in your mind:

DO NOT EVER TOSS AN ANTENNA WIRE OVER THE AC POWER LINES!

If you plan to do this trick, then have your family prepare for some solemn praying, sad music, and slow marching, for your funeral is imminent. This action kills radio hobbyists every year. There is no safe way to do it, so do not. Even if both the antenna wire and the power line are insulated, the forces of erecting the antenna will slice through both insulations! When that happens, the wires will come into contact with each other and with you – with disastrous consequences: human bodies make lousy fuses.

The next matter is like unto the first: do not erect your antenna where it can fall onto the power lines, or other people's property, either during erection or as a result of a structural failure. Falling onto the electrical lines can make the antenna deadly, and falling onto your neighbor's property is just plain rude (and might get you sued).

Finally, do not install an antenna without assistance. I know they are lightweight, and all that, but they have a remarkable 'sail area,' and even very weak winds will increase their apparent weight. I've suffered injury myself (wrenched back) by working alone, and a friend of mine broke a pelvis and a leg falling off a roof while trying to install a television antenna.

CHAPTER 4

Marconi and other unbalanced antennas

Marconi antennas are those that are *fed unbalanced with respect to ground* (Figure 4.1). In the case of a transmitter, one side of the output network is grounded, while the other side is connected to the radiator element of the antenna. The length of the radiator element varies with type, and determines many of the operating properties of the antenna. The direction of the radia-

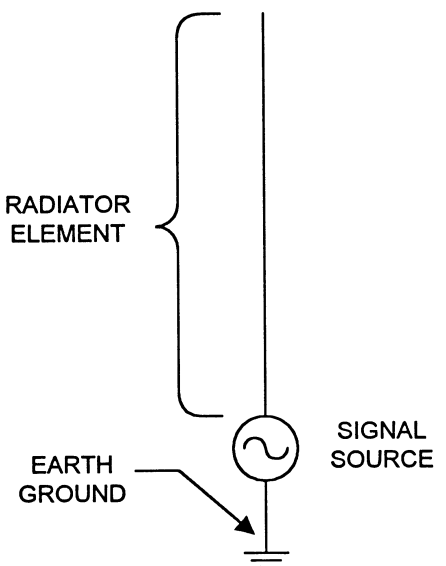


FIGURE 4.1

tor element also varies. It may be vertical as shown (indeed, vertical antennas are a species of Marconi antenna), or at a slant. In some cases, the Marconi antenna is connected directly to the rig or receiver, but in other cases an antenna tuning network is needed.

BASIC RECEIVING 'TEE' ANTENNA

Although most of the antennas in this chapter can be used on either receive or transmit, this one (Figure 4.2) is recommended for receive use only unless an antenna tuner is provided. This antenna is not particularly well thought-of, but is the antenna of choice for many first-time receiver owners, or those who live in a difficult antenna erection venue.

The radiator element is a 10–40 m long (more or less) piece of wire, preferably in the 12 AWG (≈14 SWG), 14 AWG (≈16 SWG), or 16 AWG (≈19 SWG) size. Smaller sizes can be used, but they are not recommended for structural reasons (they break and fall down too easily).

THE HALF-LAMBDA-'TEE' ANTENNA

The antenna in Figure 4.3 is called the half-lambda-'tee' because it is one-half wavelength long (i.e. $\lambda/2$) and is a 'tee' antenna. The 'tee' is seen by the fact that a downlead section is provided, and it comes away from the radiator element at a 90° angle. The length of this downlead section is quarter wavelength. This length differs from the top-hat tee antenna used by many

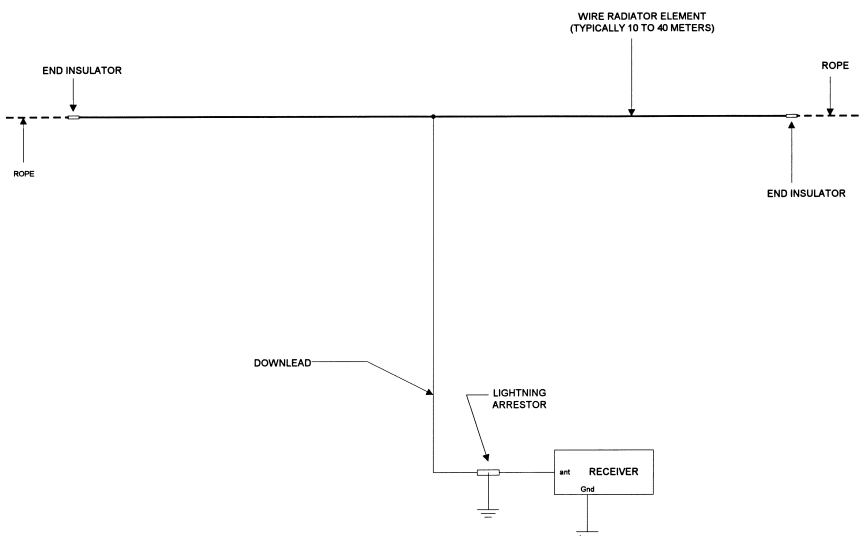


FIGURE 4.2

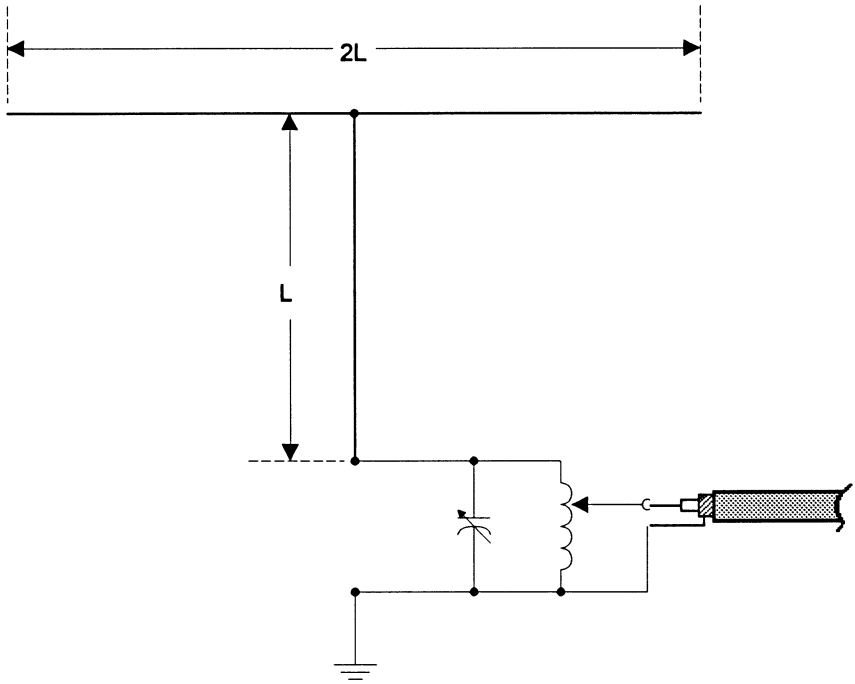


FIGURE 4.3

short-wave listeners where the radiator is half wavelength or longer and the downlead is a random length of wire. The lengths of these elements are

$$2L = \frac{143}{F_{\text{MHz}}} \text{ meters}$$

and

$$L = \frac{71.5}{F_{\text{MHz}}} \text{ meters}$$

These lengths make the antenna resonant. Examples of the proper lengths for various bands are:

Band	L	$2L$
60 m (5 MHz)	14.30 m	28.60 m
31 m (9.75 MHz)	7.33 m	14.66 m
20 m (14.2 MHz)	5.03 m	10.06 m

An antenna tuner is required for the half-lambda-'tee' antenna. Both the inductor and the capacitor should have reactances of 450 ohm, i.e.

$X_L = 450$ ohms and $X_C = 450$ ohms. Typical values for these components are:

Band	*C*	*L*
60 m (5 MHz)	65 pF	16.0 μH
31 m (9.75 MHz)	33 pF	8.2 μH
20 m (14.2 MHz)	23 pF	5.6 μH

For operation over a wider band, try using variable capacitor and/or inductor elements.

THE 'QUICK 'N' DIRTY' TWIN-LEAD MARCONI ANTENNA

The antenna shown in Figure 4.4 is one of the first that I ever used (when a mentor, MacIvor Parker, W4II, gave me a roll of twin-lead wire). I ran this antenna out of a basement window, and attached the top end to an old pine tree, and the middle supported by a 5×10 cm piece of construction lumber nailed rather unceremoniously to a tool shed on the back of the house.

Basically a quarter-wavelength 'round robin' Marconi, it consists of a 300 or 450 ohm twin-lead transmission line with a length (in meters) of $71.5/F_{MHz}$. Thus, for the 60 m (5 MHz) band the length would be about 14.3 m long overall. The length of the coaxial cable feedline is not critical. The 52 ohm coaxial cable feedline is connected such that the center conductor goes to one conductor of the twin-lead, and the shield goes to the other.

This antenna is basically a species of Marconi antenna, so requires a good ground in order to work properly. At the minimum, one of the longer copper or copper-clad steel ground rods should be driven into the earth at

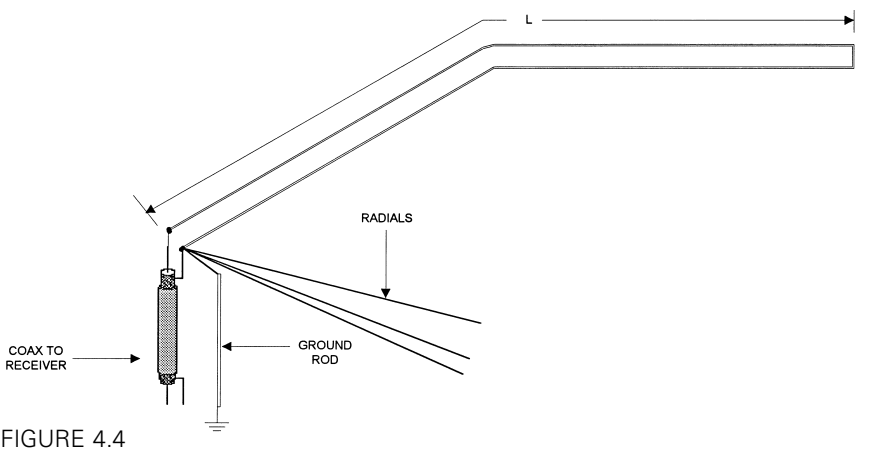


FIGURE 4.4

or very near the feedpoint. The best method is to use both a ground rod and a system of two to eight quarter-wavelength radials as a counterpoise ground.

The 'quick 'n' dirty' (or 'round robin' as my mentor called it) can be installed at an angle (as shown), although I suspect best performance occurs when the wire is straight. In the version that I had, there was only a small voltage standing wave ratio (VSWR) with one bend that put the vertical and horizontal segments at about a 120° angle.

Note the far end of the twin-lead wire (i.e. the end away from the coaxial cable). The two conductors of the twin lead are *shorted together*. I have encountered some confusion over this matter, after all a short is a short, isn't it? No, because this antenna is a quarter wavelength long, so the impedance reflected to the feedpoint by a short at the end is very high.

This antenna works better than single-conductor Marconi antennas because regular Marconi antennas suffer from ground losses. By supplying a return line, the radiation resistance is raised from 10 or 15 ohms to something of the order of 40–50 ohms – which makes it a good match to 52 ohm coaxial cable.

THE SWALLOW TAIL ANTENNA

Figure 4.5 shows two variants of the swallow tail multiband antenna. The radiator elements in both Figures 4.5A and 4.5B are cut to specific frequency bands, and are quarter wavelength ($L_{\text{meters}} = 75/F_{\text{MHz}}$). As many as needed to cover the bands of interest may be used, provided that they do not bear a 3:1 frequency ratio. The reason for the constraint is that for all antennas but the resonant one the impedances are so high that connecting them in parallel with each other does not affect the overall feedpoint impedance. However, at the third harmonic, the impedance again drops low and will load down the impedance seen by the transmission line. This is another situation where the results are more profound for amateur radio transmitters than for shortwave receivers, but it is a good idea to avoid it if possible. Besides, the antenna is actually resonant on its third harmonic, so you lose nothing.

The version in Figure 4.5A uses a pair of insulating masts, or support structures (roof of a house, tree, etc.) with a rope stretched between them. The quarter-wavelength resonant wires are spread out along the length of the rope, spaced approximately evenly.

The version shown in Figure 4.5B uses a large wooden cross-like structure. The antenna wires are connected to the cross-piece at the top end, and to the coaxial cable at the bottom end. This antenna apparently worked well for a fellow who wrote to me recently. He lives in a townhouse community that has a 'homeowners' association' of nit-picking little dictators who like

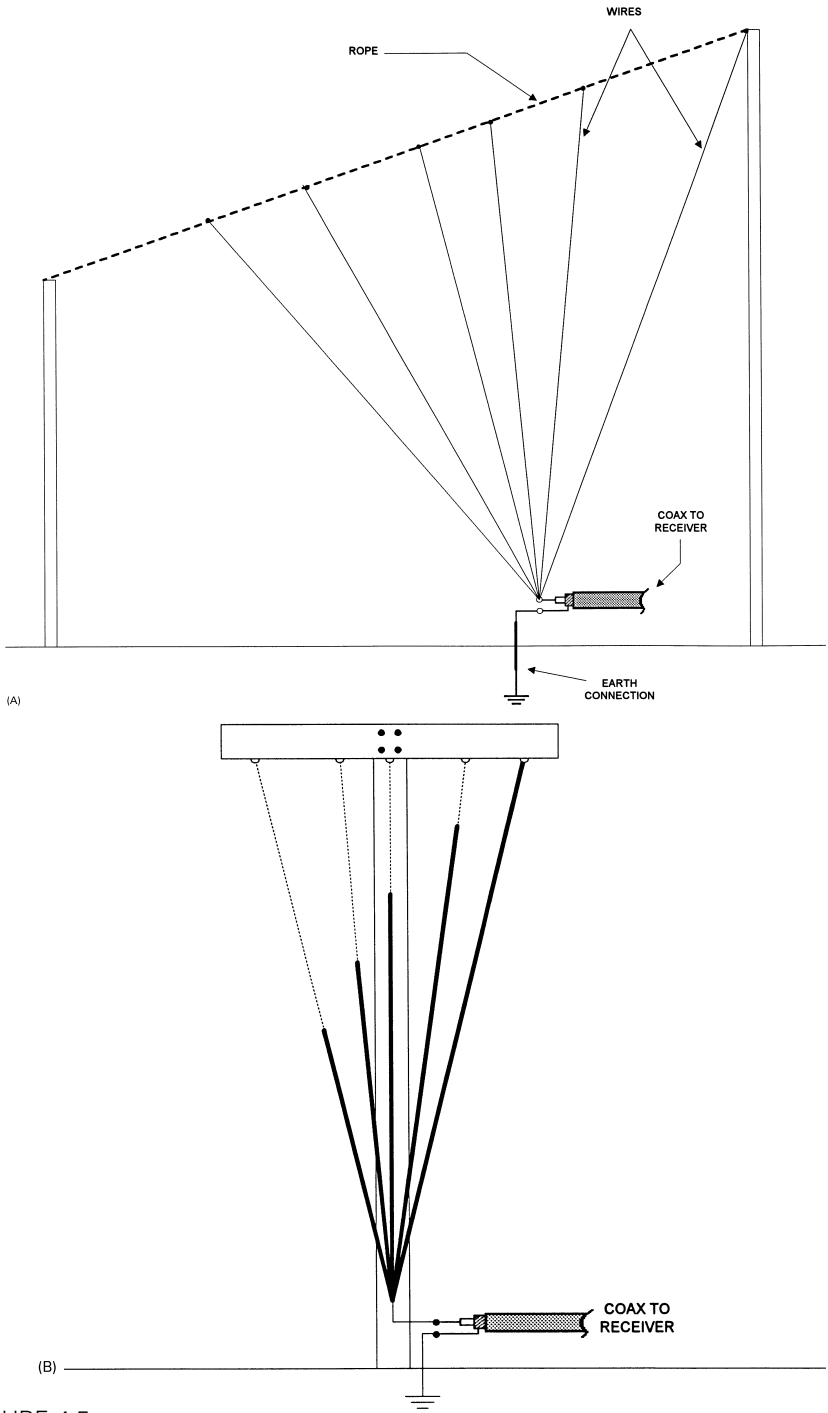

WIRES
ROPE
COAX TO
RECEIVER
EARTH
CONNECTION
(A)
COAX TO
RECEIVER
(B)

FIGURE 4.5

to tell people what they may do with their houses. One of the rules is 'no outdoor antennas' of any sort. He erected a mast in his rear garden shaped like a ship's mast and yardarm, and then convinced the busy-bodies snooping for 'the committee' (dread!) that it was somehow nautical, not radio.

RANDOM LENGTH MARCONI

There are several antennas that qualify as 'all-time favorites' with both ham operators and short-wave listeners. In the top three or four is the random-length wire antenna. This form of antenna consists of whatever length of wire that you happen to have (although there are reasonable limits!) run from the rig or receiver to a convenient elevated support (e.g. a tree, mast, or roof of a building). Figure 4.6 shows this most basic form of Marconi antenna.

Although a receiver owner can connect the random-length radiator wire to the antenna input of the receiver, the transmitter operator most certainly requires an antenna tuning unit (ATU). Even receiver operators will find this antenna a better performer when an ATU is used between the coaxial cable to the receiver and the radiator wire.

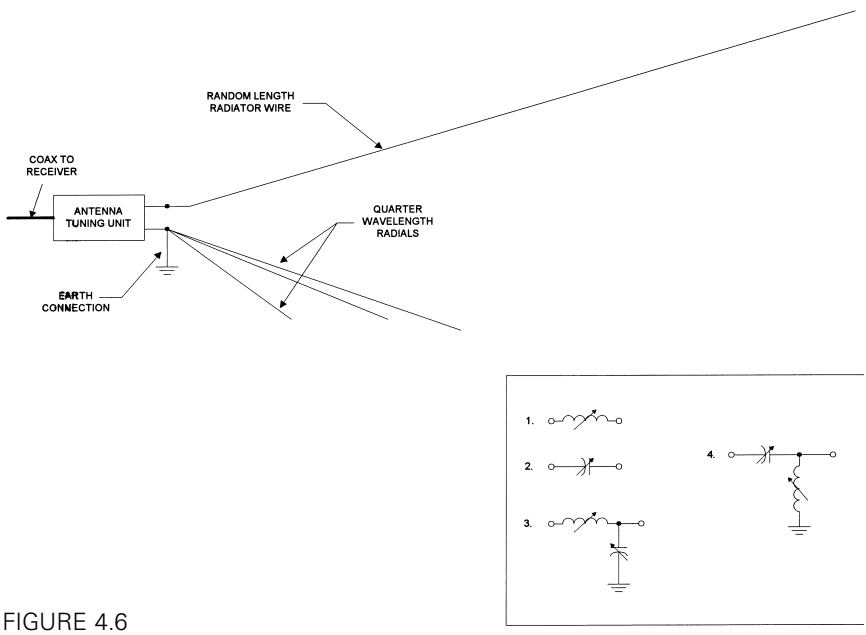


FIGURE 4.6

One requirement of this antenna is that it requires a good ground. It is essential that either an earth ground or a counterpoise ground (i.e. a system of radials) be provided. Methods of providing a good grounding system are described in Chapter 13.

There are several conditions for this form of antenna: an antenna that is shorter than a quarter wavelength, an antenna that is a quarter wavelength, and longer than a quarter wavelength. The inset in Figure 4.6 shows several options for the antenna tuning unit. At position 1 is a variable inductor. This is used when the antenna is quite short compared to a quarter wavelength. At position 2 is a variable capacitor. It is used when the antenna is longer than a quarter wavelength. When the antenna is exactly a quarter wavelength, then it may be possible to eliminate the antenna tuner. It is also possible to use the L-section couplers at positions 3 and 4. These ATUs combine the inductor and capacitor. These are used to match the feedpoint impedance to the 52 ohm coaxial cable.

Many commercial L-section couplers are designed with either switching or shorting straps that allow any of the combinations in the inset to Figure 4.6 to be realized by making some simple changes.

Receiver operators who use this antenna are well advised to use an ATU, just like transmitter operators. But there is at least one difference: you should be able to switch the ATU out of the system when you are receiving a frequency outside the limits of the tuner.

THE WINDOM ANTENNA

One of the first ham radio transmitting antennas that I ever used was not my own, but rather the antenna at my high school station in Arlington, Virginia (K4BGA, Washington-Lee High School). Although I had owned receive antennas, the K4BGA club station was one of the first I had operated. It used an HF Windom antenna running across the roof of the industrial arts building. The Windom antenna (Figure 4.7) has been around since the 1920s. Although Mr Loren Windom is credited with the design, there were actually a number of contributors. Co-workers with Windom at the University of Illinois were John Byrne, E. F. Brooke, and W. L. Everett, and they are properly co-credited. The designation of Windom as the inventor was probably due to the publication of the idea (credited to Windom) in the July 1926 issue of *QST* magazine. Additional (later) contributions were rendered by G2BI and GM3IAA (Jim MacIntosh). We will continue the tradition of crediting Loren Windom, with the understanding that others also contributed to this antenna design.

The Windom is a roughly half-wavelength antenna that will also work on even harmonics of the fundamental frequency. The basic premise is that the antenna radiation resistance varies from about 50 ohms to about 5000

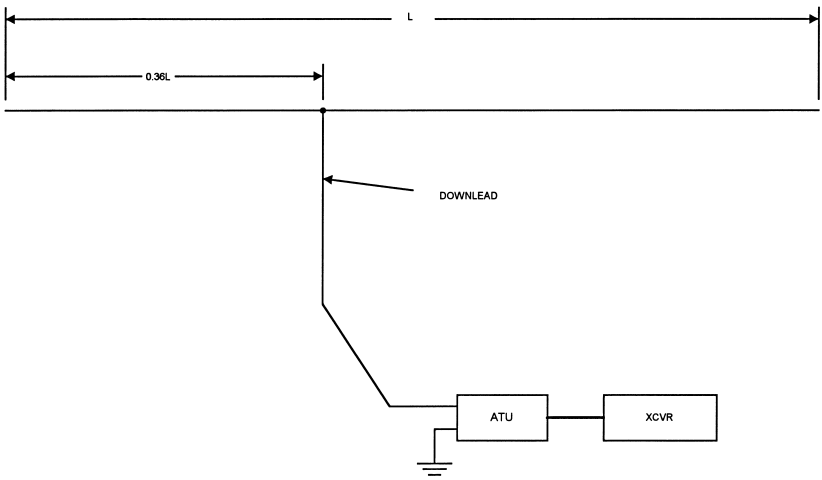


FIGURE 4.7

ohms, depending upon the selected feedpoint. When fed in the exact center, a current node, the feedpoint impedance will be 50 ohms; similarly, end-feeding the antenna finds a feedpoint impedance of about 5000 ohms. In Figure 4.7 the feedpoint is tapped away from the center at a point that is about one-third ($0.36L$) the way from one end, at a point where the impedance is about 600 ohms.

The feedline for the basic Windom of Figure 4.7 is an insulated length of wire. Of course, the size of the wire depends on the power level, but I suspect that No. 14 insulated stranded wire will do for most people who run less than 200 W of power (indeed, I would not like to use a Windom at high-power levels because of the 'radio frequency (RF) in the shack' problem).

The Windom antenna works well – but with some serious caveats. For example, the antenna has a tendency to put 'RF in the shack' because of the fact that it is voltage fed. This is why Windom antenna users (along with those using the random-length Marconi) get little 'nips' of RF shock when touching the transmitter chassis, or putting a lip on the microphone. Second, there is some significant radiation loss from the feedline. Finally, the antenna works poorly on odd harmonics of the fundamental frequency.

The antenna tuning unit can be either a parallel resonant, link-coupled, LC tank circuit or a π-network (i.e. high-impedance end toward the antenna).

Note that a good ground should be used with this antenna (note the ground connection at the output of the antenna tuning unit). This basically means (for most people) a 2–2.5 m ground rod, or a system of radials (see earlier discussion for random-length wire antennas).

A reasonable compromise Windom, which reduces feedline radiation losses, is shown in Figure 4.8. In this antenna a 4:1 BALUN transformer is placed at the feedpoint, and this in turn is connected to 75 ohm coaxial cable. It is necessary to ensure that the feedpoint is at a current node, so it is placed a quarter wavelength from one end. Also, the antenna on the other side of the BALUN transformer should have a length that is an odd multiple of a quarter wavelength. In the case shown it is three-quarter wavelength long.

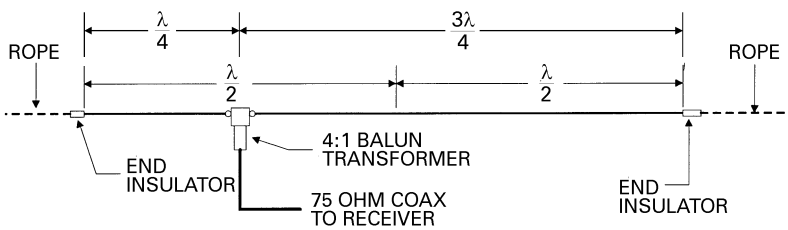


FIGURE 4.8

QUARTER WAVELENGTH VERTICAL

Perhaps the most classic form of Marconi antenna is the quarter wavelength vertical (Figure 4.9). It consists of a vertical radiator element that is quarter wavelength long, and a system of quarter wavelength radials for the ground system. The length of the radiator element is found from:

$$L_{\text{meters}} = \frac{75}{F_{\text{MHz}}}$$

where L_{meters} is the length in meters and F_{MHz} is the frequency in megahertz.

The radiator element can be made of pipe or wire. The former case, the pipe can be copper or aluminum, although I suspect that aluminum will be the choice of most readers. In the case of a wire radiator, it will be necessary to support the wire up top, so some form of mast or support is necessary. Make the mast or support out of an insulating material such as wood.

The radials are somewhat optional if you use it with a 'good ground'. But in the absence of a 'good ground', I would use at least four radials and as many as sixteen. Above sixteen the benefit of adding further radials drops off substantially.

Radials should be buried if the vertical is ground mounted (that is to prevent lawsuits over pedestrians tripping over them). In the case where the

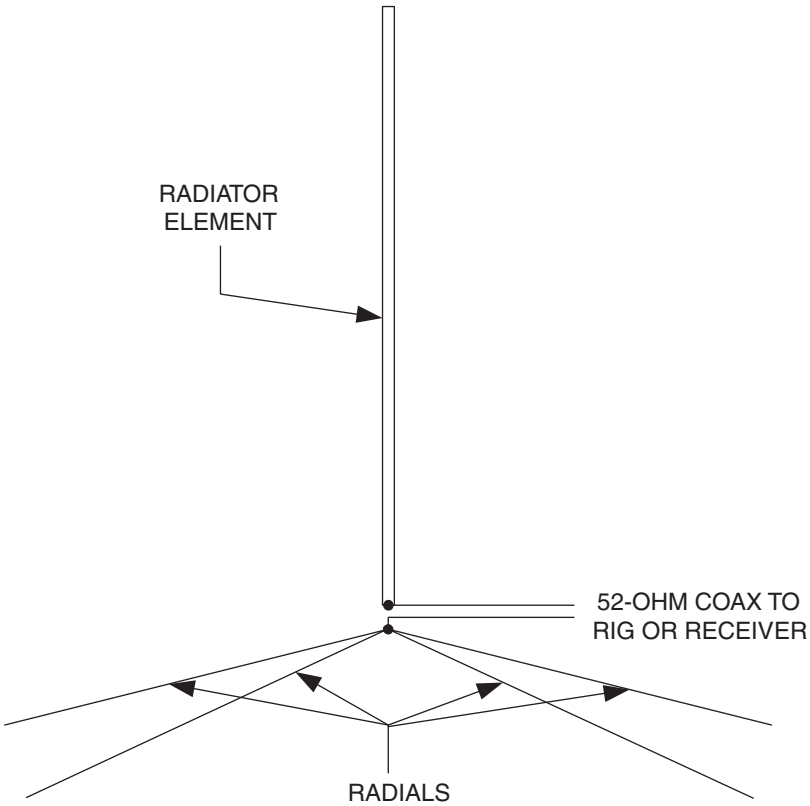


FIGURE 4.9

vertical is mounted on a mast, off ground, the radials are not optional – they are mandatory. In that case, use four to sixteen radials arranged in a circular pattern around the base of the radiator element.

Connect the system so that the radiator element is connected to the coaxial cable center conductor, and the radials are connected to the shield of the coaxial cable. Because the feedpoint impedance of the vertical antenna is 37 ohms, use of 52-ohm coax will result in a VSWR of only 50/37 = 1.35:1. In the case where the vertical has a substantially lower impedance (as low as 2 ohms!), use a broadband transformer between the coaxial cable and the antenna.

FOLDED MARCONI 'TEE' ANTENNA

The antenna shown in Figure 4.10 is popular for receiving and transmitting on the lower bands (< 7 MHz) when space for antennas is a bit limited.

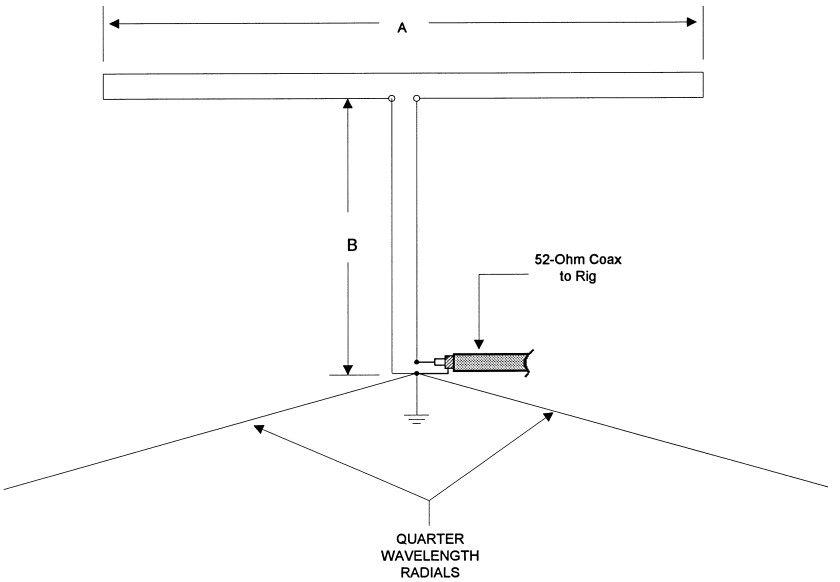


FIGURE 4.10

Although it can still take a lot of space at the lowest frequencies, it is quite a bit shorter than a half-wavelength dipole for the same frequencies.

Two pieces of 300 or 450 ohm twin-lead are used to make this antenna (*A* and *B* in Figure 4.10). The *A*-section is the main radiator element, and it has a length of

$$A_{\text{meters}} = \frac{82}{F_{\text{MHz}}} \text{ meters}$$

Note that section *A* is built like a folded dipole. At the ends of the twin-lead the two conductors are shorted together. At the center point of section *A* one of the two conductors is cut to accept the two conductors of the twin-lead used for section *B*.

Section *B* is vertical, and should come away from section *A* at a right angle. It has a length that is similar to that of *A* except that it is reduced by the velocity factor (VF) of the transmission line:

$$B_{\text{meters}} = \frac{82(\text{VF})}{F_{\text{MHz}}} \text{ meters}$$

For ordinary 300 ohm twin-lead, the length of *B* is $0.82A$.

Note that the conductors making up *A* and *B* form a continuous loop of wire. The coaxial cable is connected such that the shield goes to one lead of *B* and the inner conductor goes to the other conductor. At the top of *B* the two conductors are connected to either side of the cut in section *A*.

As with the other Marconi antennas in this chapter, it is imperative that a good ground be used with the 'tee' Marconi. Otherwise, losses will be too great, and performance will suffer considerably.

Figure 4.11 shows the end-fed Zepp antenna. This antenna has a radiator wire that is half wavelength long at the lowest frequency of operation. It will work on harmonics of that frequency as well as the frequency itself. It will also work on other frequencies if a high VSWR can be tolerated. It is fed by 600-ohm parallel feedline and an antenna tuning unit (ATU). The ATU is used to tune the VSWR to 1:1 at the frequency of operation.

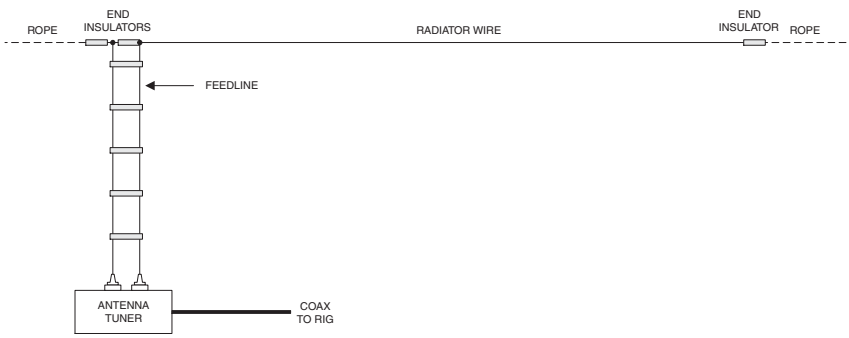


FIGURE 4.11

EWE ANTENNA

The EWE antenna emerged recently as one solution to the low-noise low-band antenna problem. Figure 4.12 shows the basic EWE antenna. It consists of two vertical sections (labeled L1) and a horizontal section (L2). The EWE looks superficially like a Beverage antenna, but it isn't. Like the Beverage it is erected about L1 = 3 meters above the Earth's surface. Unlike the Beverage, it is only L2 = 6.5 meters long at 3.5 MHz. Those dimensions make it affordable for most people.

The far end segment is terminated in an 850 ohm resistor. This resistor should be a carbon composition or metal film resistor, and never a wire-wound resistor.

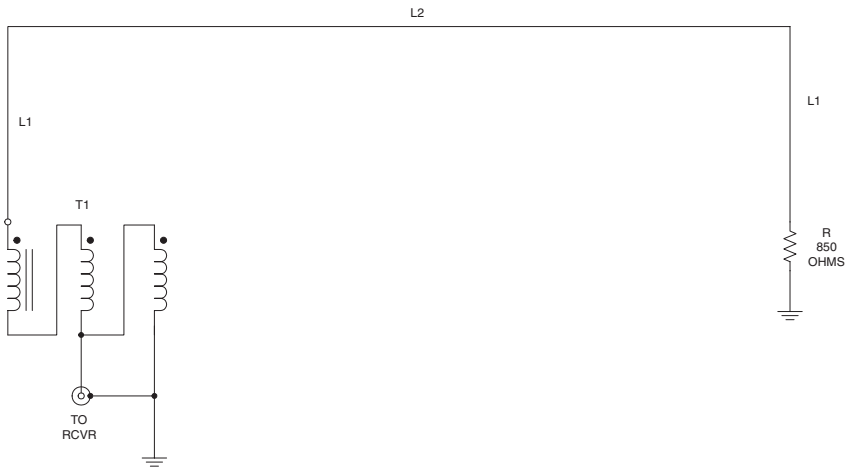


FIGURE 4.12

The receiver end must be matched to the receiver's 50 Ω antenna input impedance. Transformer T1 is provided for this purpose. It has a turns ratio of 3:1 to provide the 9:1 impedance ratio required to match the 450 Ω antenna impedance to the 50 Ω receiver impedance. A powdered iron toroid core made of -2, -6 or -15 material will be sufficient. A suitable transformer can be made using a T-50-15 (red/white) core. Use about 20 turns of any size enameled wire.

The azimuthal and elevation patterns for the Koontz EWE antenna are shown in Figures 4.13, 4.14, 4.15 and 4.16. These patterns were simulated from the *Nec-WIN Basic* software available from Nittany-Scientific. The patterns in Figures 4.13 and 4.14 are based on the Sommerfield-Norton standard ground model, with the azimuth being seen in Figure 4.13 and the elevation in Figure 4.14. The same types of pattern are seen for a 'real' ground based on suburban soil and are shown in Figures 4.15 (azimuth) and 4.16 (elevation).

REVERSIBLE EWE

The EWE antenna can be made reversible by using a system such as Figure 4.17. The feedpoint and termination circuits are co-located at the receiver. Transformer T1, coil L1, resistor R1 and DPDT switch S1 are installed in a shielded metal box. The outputs of the box (i.e. center terminals of the DPDT switch) are connected to the bases of the vertical (L1) sections of the EWE antenna. According to one source, the simple resistive termination

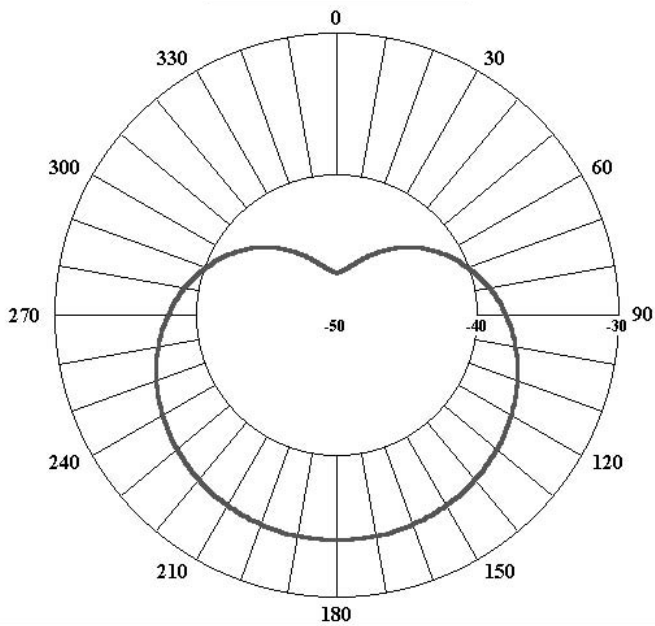


FIGURE 4.13 Low-Noise RX Antenna

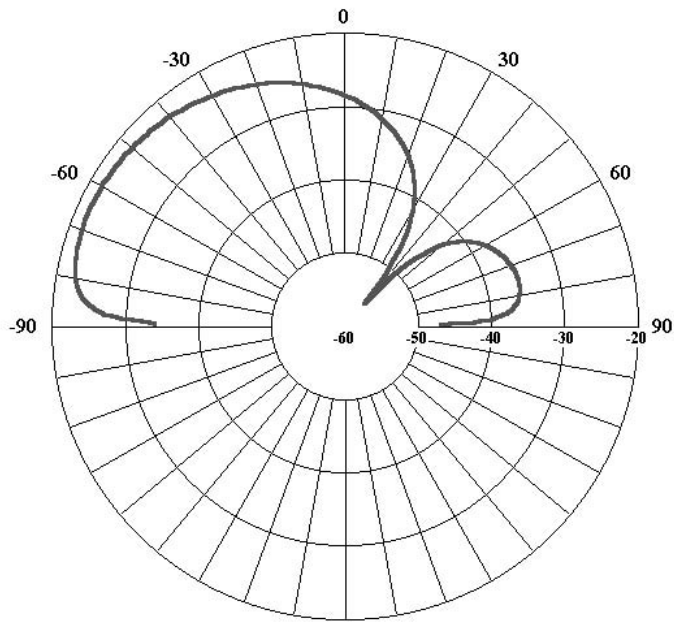


FIGURE 4.14 Low-Noise RX Antenna

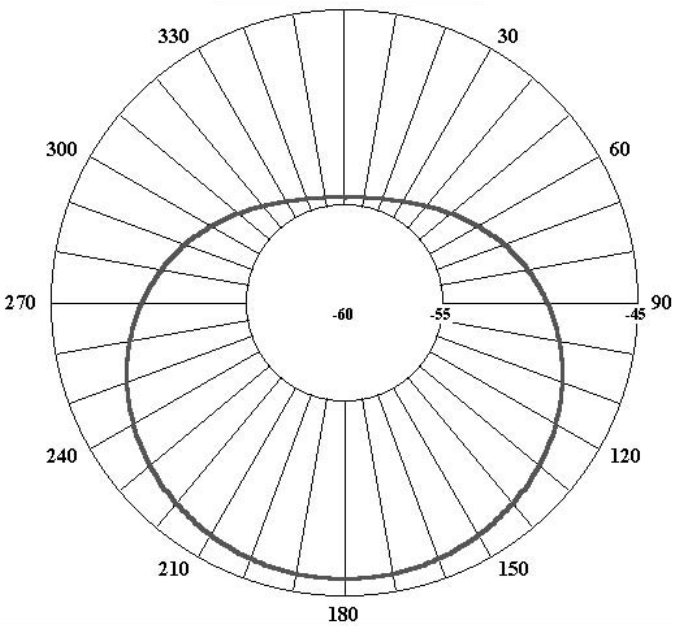


FIGURE 4.15 Low-Noise RX Antenna Rocky Soil

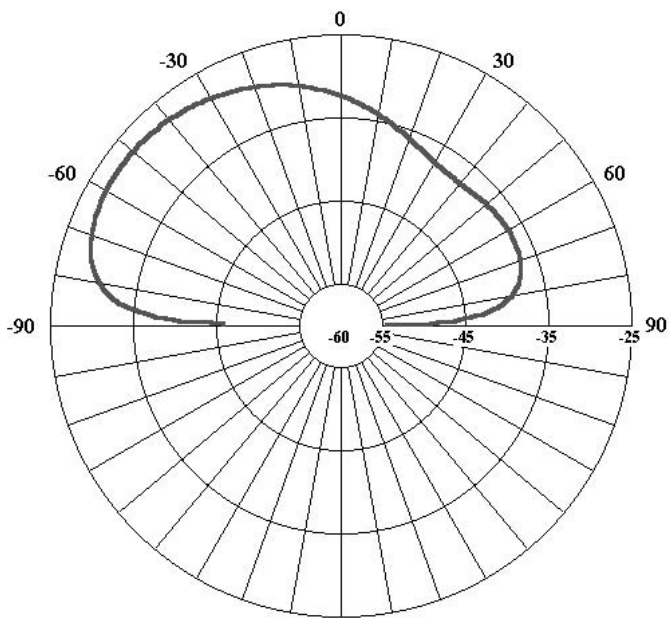


FIGURE 4.16 Low-Noise RX Antenna Rocky Soil

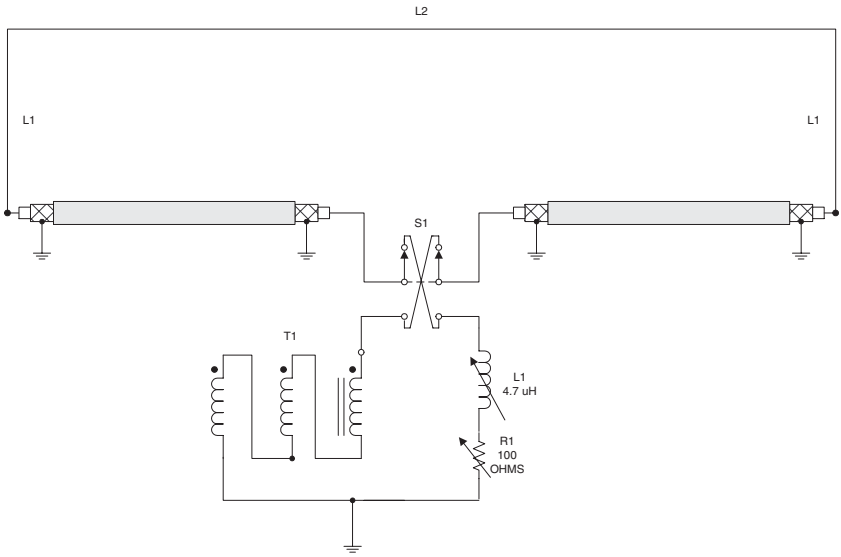


FIGURE 4.17

was not sufficient, so they added an inductive reactance in series with a resistance. This is the method used on Beverage antennas to make a steerable null, and that effect is seen on the EWE as well.

DUAL EWE ANTENNA

Figure 4.18A shows a modification of the EWE antenna that permits switchable bi-directionality. Four EWE antennas are arranged in north–south (N–S) and East–West (E–W) directions. The feedpoints (A, B, C and D) are connected to a switch circuit such as shown in Figure 4.18B. The directivity of the antenna is controlled by opening and closing the four switches (S1–S4).

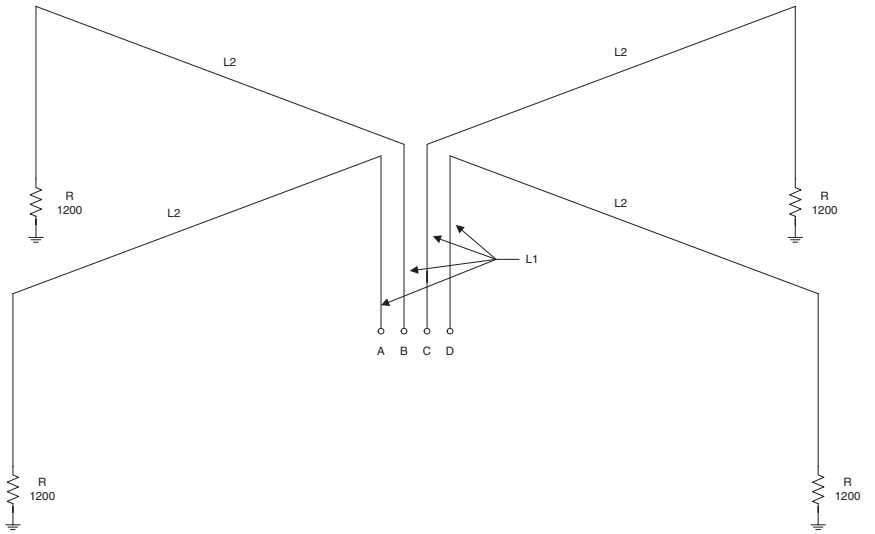


FIGURE 4.18A

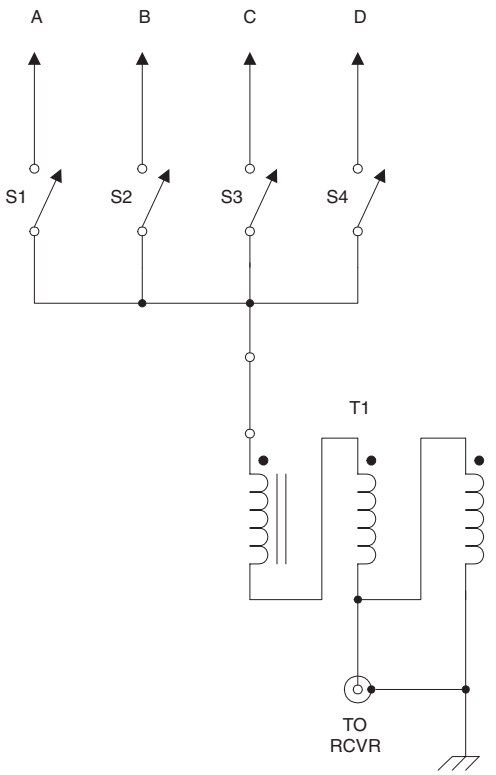


FIGURE 4.18B

CHAPTER 5

Doublets, dipoles, and other Hertzian antennas

A **doublet** antenna is any of several forms of **dipoles** (*di* – 'two') that are balanced with respect to ground. In other words, neither side of the antenna feedline is grounded. These antennas are fed in the center, which point will be either a current node or voltage node depending on the design of the particular antenna. Perhaps the most common form of dipole is the half-wavelength ($\lambda/2$) center-fed dipole, which can be fed with ordinary 75 ohm coaxial cable. It is so popular that when people use the word 'dipole' in an unqualified sense, it is the $\lambda/2$ center-fed version that is meant. Although dipoles can be made of either wire or tubing, the most common practice is to make them out of wire, especially antennas at the lower end of the high-frequency (HF) spectrum. Above about 14 MHz, however, it becomes increasingly practical to make these antennas of aluminum tubing.

DIPOLE RADIATION PATTERNS

When one uses the term 'dipole' in the common sense it is usually understood that the antenna being discussed is the half-wavelength doublet. One reason for this practice is the widespread popularity of that particular dipole. But the practice is actually erroneous because the dipole can be any size up to about two wavelengths (although some would argue that any balanced antenna, even longer than two wavelengths, is a dipole, I demur on grounds that longer antennas begin acting like the **long-wire** antenna and its relatives). Figure 5.1 shows the effect of different wavelength dipoles on the azimuthal radiation pattern of the antenna.

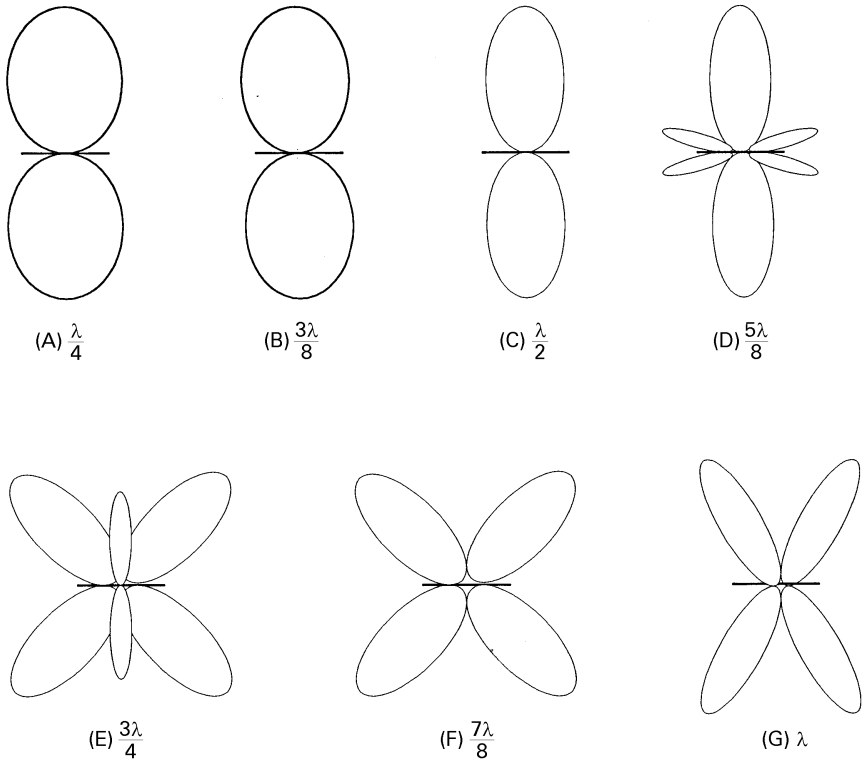


FIGURE 5.1

At $\lambda/4$, $3\lambda/8$, and $\lambda/2$ the azimuthal pattern is the classic figure '8' that is commonly associated with dipoles. The principal difference in these patterns is the particular shape (width, etc.) of the pattern, which varies with antenna size. At $5\lambda/8$ wavelength, the figure '8' persists, but becomes narrower and minor lobes begin to appear. To a receiver, these sidelobes represent responses in directions that may not be desired, and for transmitters it represents wasted energy. At $3\lambda/4$, the pattern blossoms into a four-lobe 'clover leaf' with significant minor lobes. This pattern is seen in ham radio when a half-wavelength 40 m (7 MHz) dipole is used at 15 m (21 MHz). The same physical length that is $\lambda/2$ at 7 MHz is $3\lambda/4$ at 21 Mhz. At $7\lambda/8$ wavelength the minor lobes disappear, and at 1λ the pattern thins out a bit but retains the clover leaf shape.

HALF-WAVELENGTH DIPOLE

It is often the case that a half-wavelength dipole is the first antenna that a new ham operator or short-wave listener will put up. These antennas are low

cost, easy to build, and have a delightful tendency to work well with only a little muss and fuss. They are thus well suited to the newcomer.

Figure 5.2A shows the basic half-wavelength dipole. It consists of a half-wavelength radiator ('B') that is cut into two sections ('A') which are each a quarter-wavelength long. The feedpoint is the middle of the half-wavelength, and for this size dipole the feedpoint is a current node. The feedpoint impedance is therefore a minimum, and the dipole makes a good match to 75 ohm coaxial cable.

It is commonly assumed that the feedpoint impedance of the dipole is 73 ohms, but that is a nominal value which is only true at certain heights above the ground. As can be seen in Figure 5.2B, the actual impedance is a function of the height of the dipole, and at a certain height it converges to 73 ohms. At other heights it can vary from a very low impedance (near zero ohms) up to about 110 or 120 ohms.

The reason for the impedance situation is shown in Figure 5.2C. This graph shows the distribution of voltage and current along the length of the antenna element. Note that the current peaks at the feedpoint and drops to a zero at the ends. The voltage, on the other hand, peaks at the two ends and drops to a minimum at the feedpoint. One of the principal differences between different dipole designs is the issue of whether they are voltage fed or current fed.

The element lengths (in meters) of the half-wavelength dipole would normally be found from $150/F_{\mathrm{MHz}}$ for the overall length, and $75/F_{\mathrm{MHz}}$ for each quarter-wavelength segment. However, because of the velocity factor effects of the length–diameter ratio (which is high for wire HF antennas), and the capactive end effects, a small foreshortening occurs, making the actual lengths closer to

$$B_{\mathrm{meters}} = \frac{143}{F_{\mathrm{MHz}}} \text{ meters}$$

$$A_{\mathrm{meters}} = \frac{71.5}{F_{\mathrm{MHz}}} \text{ meters}$$

These lengths are still only approximate, although in many cases they will be right on the money. There is nonetheless a possibility of needing to lengthen or shorten the antenna if an exact resonant frequency is needed. This is done by either noting the voltage standing wave ratio (VSWR) for the minima being at the desired resonant frequency, or by the feedpoint impedance being close to the design value.

The standard dipole is fed in the center with coaxial cable. For installations really on the cheap, ordinary lamp cord (alternating current wiring) can be used, as it looks like 75–100 ohm parallel transmission line to the antenna. However, except for receiver and the lowest power transmit

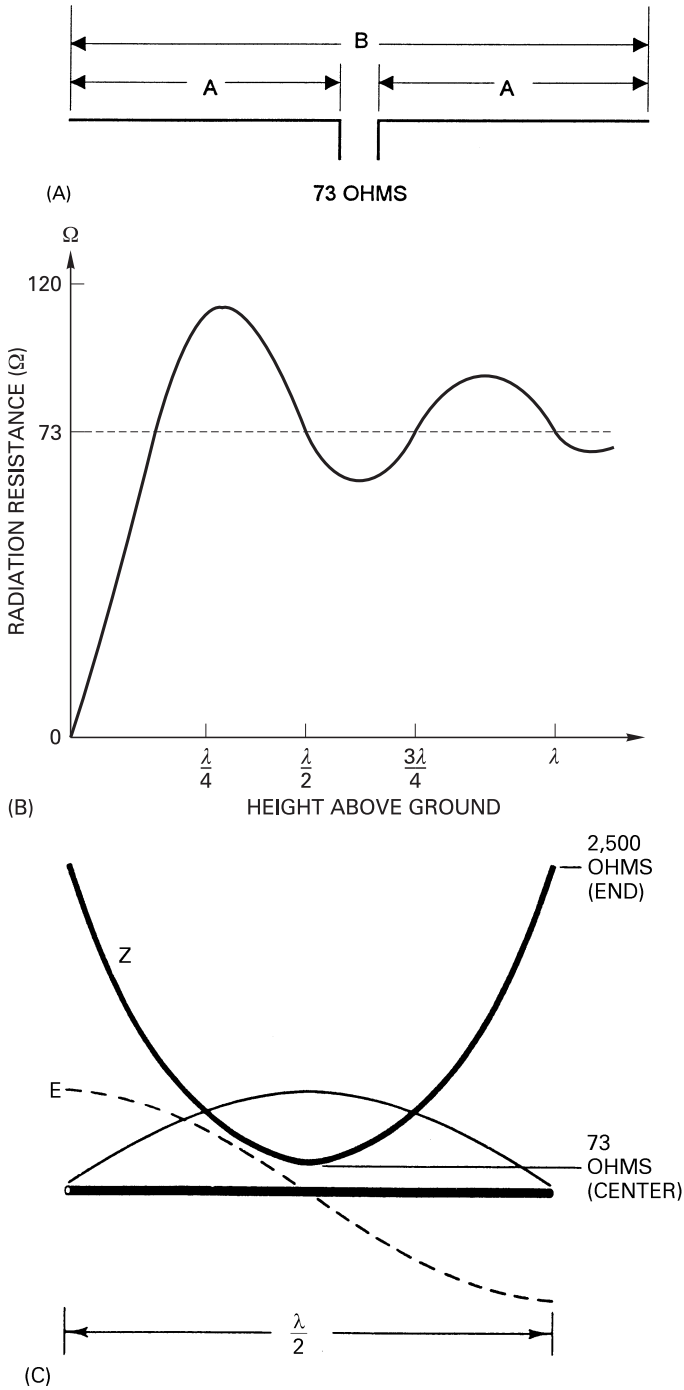

B
A
A
(A)
73 OHMS
Ω
120
73
0
RADIATION RESISTANCE (Ω)
λ/4
λ/2
3λ/4
λ
(B)
HEIGHT ABOVE GROUND
2,500
OHMS
(END)
Z
E
73
OHMS
(CENTER)
λ/2
(C)

FIGURE 5.2

operations, it is definitely not recommended. Even receiver operators will fare better with coaxial cable because it has superior noise immunity.

In the usual installation, the dipole will have a center insulator to support the middle ends of the two sections, and to provide the connection spot for the transmission line. The outer ends of the two sections will be supported by end insulators and rope or heavy twine tied to a mast, roof top, or convenient tree. Antenna and radio shops sell special center insulators that provide strain relieved coaxial connections for the transmission line and solder points for the antenna elements.

To BALUN or not to BALUN, that is the question

The BALUN transformer is used to convert between balanced and unbalanced loads. In the context of the dipole, the feedpoint is balanced with respect to the ground, while the transmitter output is unbalanced with respect to the ground. What this means in plain language is that an unbalanced load has one side grounded, while a balanced load is not grounded. Some BALUN transformers also provide impedance transformation, with 4:1 being the most commonly seen. This means that a load of R across the output ports of the transformer is reflected back to the input so that 'looking into' the transformer it will seem like a load of $R/4$. This is the transformer used to reduce 300 ohms from television-style twin-lead to 75 ohms for coaxial cable.

It is common practice to use a 1:1 BALUN transformer at the feedpoint of the half-wavelength dipole antenna (Figure 5.3), even though it provides no impedance transformation (and none is needed). The BALUN transformer balances the load, causing equal but opposite phase currents to flow in the two conductors, and thereby reduces radiation from the transmission line. The effect of the feedline radiation is to distort the radiation pattern. Figure 5.4 shows the effect on the pattern, at least in simplified form. Figure 5.4A shows the 'normal' pattern, which is what we saw in Figure 5.1C. But radiation from the feedline, and unbalanced current flow, tends to cause the nulls to fill in and the main lobes to reduce a bit (Figure 5.4B). In some cases, the afflicted pattern is a lot more distorted than shown in Figure 5.4B, which is a highly simplified case. When a BALUN transformer is used, however, the resultant pattern will look a lot nearer the ideal than the distorted version, and the dipole will perform as expected.

The use of the BALUN in the half-wavelength dipole is one of those controversial things that radio enthusiasts debate seemingly endlessly, but there is no real controversy. The simple fact is that half-wavelength dipoles work better in practical situations when a 1:1 BALUN is used to interface the coaxial cable transmission line to the antenna feedpoint. The only excuse to not use a BALUN, in my not so humble opinion, is the economic one.

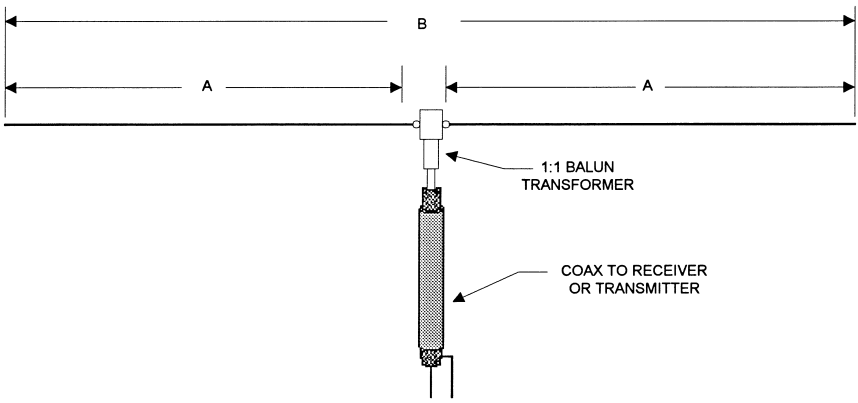


FIGURE 5.3

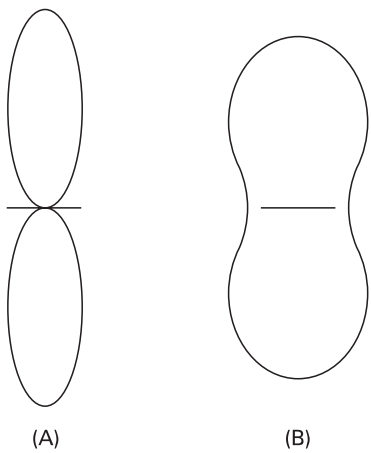


FIGURE 5.4

There were, after all, periods in my life when even so low a priced item as an HF BALUN was very dear indeed, so I used an end insulator at the feedpoint instead (once I even used a discarded toothbrush as a center insulator – and it even lasted a month or two before it broke).

MULTIBAND AND WIDE-BAND HALF-WAVELENGTH DIPOLES

The half-wavelength dipole is a resonant antenna, so it works best only on and near the center design frequency. The classic figure '8' pattern is found only at this frequency and certain lower frequencies (for which the antenna is $\lambda/4$ or $3\lambda/8$). The antenna also provides performance at odd integer

harmonics of the design frequency, but with a multilobed pattern (see Figure 5.1 again). But there are some things that can be done to accommodate the multiband interests of typical radio hobbiests. For one thing, we could put up a different dipole for each band. But even if you have the space, the installation would look like an aerial rat's nest of wires, and it requires switching the feedlines at the rig or receiver to change bands. Not a good idea, actually.

The other approach is shown in Figure 5.5: connect dipoles for different bands to the same transmission line. One can get away with this because the feedpoint impedances of the dipoles are very high except at their resonant frequency, so do not materially affect the feedpoint impedance of whichever dipole is in use at any given time. In Figure 5.5, dipole A1–A2 works at the highest band, B1–B2 works at a middle band, and C1–C2 works at the lowest-frequency band. One precaution is to ensure that none of the dipoles is cut to a frequency that is an odd integer multiple of the resonant frequency of one of the other dipoles. If, for example, you make C1–C2 cut to 7 MHz, and A1–A2 cut for 21 MHz, then the half-wavelength 21 MHz dipole A1–A2 will be in parallel with the $3\lambda/4$ wavelength 21 MHz operation of A1–A2. The low impedances will be in conflict and the pattern will be bizarre. Note, however, that $3\lambda/4$ operation means that proper selection will yield six-band operation from this 'three-band' antenna.

A certain precaution is needed when the multiband dipole is connected to a radio transmitter. Many of the HF amateur radio bands are harmonically

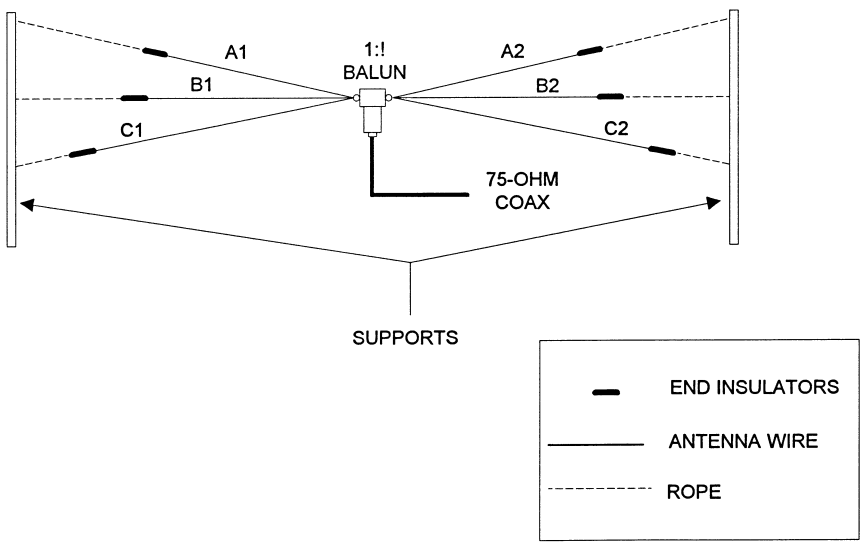


FIGURE 5.5

related. This was done, I am told by some old timers, because in the 'golden oldie' days, trapping harmonics was not as easily done as today, and amateur transmitter harmonics would fall into another amateur band, rather than onto a commercial band or public safety band. If one of the other dipoles is cut to a harmonically related band, then that harmonic looks into a resonant antenna and will radiate efficiently. If you use this type of antenna, then be certain to use a low-pass filter with a cut-off between the two bands, or use an antenna tuning unit that is not a high-pass filter (as some are), or otherwise make sure that your transmitter is harmonic free.

Figure 5.6 shows a method for producing a very wide-band dipole. This particular version was designed for use by radio science observers who listened for radio emissions from the planet Jupiter in the 16–26 MHz short-wave band. These sounds are rather strong, and can appear anytime Jupiter is above the horizon, day or night. Although a beam antenna would have been more useful in this project, the small college physics department who needed it was unable to afford a proper array, so had to make do with a simple dipole.

The idea in this design is to parallel connect three half-wavelength dipoles with overlapping frequency responses. In this particular case, the antennas were cut for 18, 21, and 24 MHz. Some amateur radio operators will use this type of antenna, but with only two dipoles, with one cut for a frequency one-third the way through the band and the other at two-thirds the way through the band. For example, in the 15 m band (21.0–21.450 MHz), the low-end antenna is cut for 21.15 MHz and the high-end antenna for 21.3 MHz.

Figure 5.7 shows how this antenna works. In Figure 5.7A we see the frequency response of a single-dipole antenna. This curve plots VSWR versus frequency. The lowest point on the VSWR curve is the resonant frequency of the antenna, and is hopefully where you designed it. Note that the VSWR climbs fairly rapidly as you get away from the resonant frequency. In Figure 5.7B we see the effect of overlapping the VSWR versus frequency curves of three antennas designed for adjacent segments of

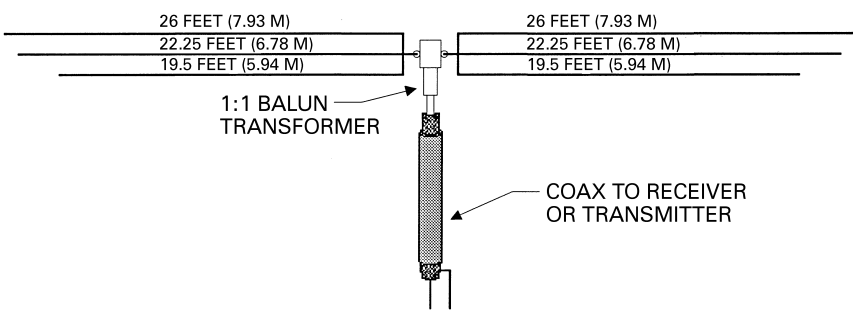


FIGURE 5.6

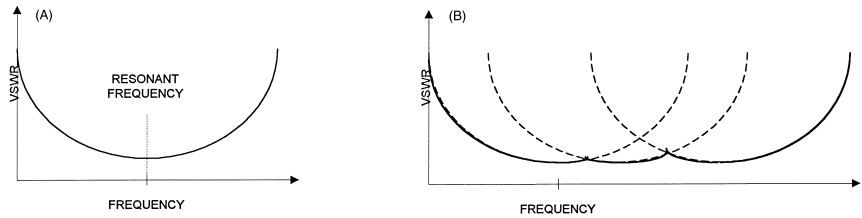


FIGURE 5.7

the band. The overall VSWR seen by the antenna is basically the lowest point on the composite curve (although there is some interaction in real antennas!).

FOLDED DIPOLES

The **folded dipole** antenna (Figure 5.8) has a wider bandwidth than the common dipole, although at the expense of a little constructional complexity. This antenna consists of a loop of wire made from a half-wavelength section of twin-lead or parallel transmission line. The most common form of folded dipole uses 300 ohm television twin-lead for the antenna element. Note that the two conductors of the twin-lead are shorted together at both ends of the radiator. The feedpoint is formed by breaking one of the two conductors at the midpoint. The feedpoint impedance of this antenna is just under 300 ohms, so it makes a good match to another piece of 300 ohm twin-lead used as a transmission line.

The overall length of the folded dipole is found from the same equation (see above) as a regular half-wavelength dipole. There may be a small amount of additional foreshortening due to the velocity factor of the twin-lead, so this length may be a tad long for the desired frequency. You can adjust to the desired frequency by trimming the same amount off of each end while monitoring the VSWR to determine the resonant frequency.

If you want to use coaxial cable to feed the folded dipole, then place a 4:1 BALUN transformer at the feedpoint. This transformer reduces the feedpoint impedance from near 300 ohms to near 75 ohms, and so makes a good match to 75 ohm coaxial cable (e.g. RG-59 or RG-11).

The folded dipole made of twin-lead can be used with moderate amounts of radio frequency (RF) power in transmitting stations, and all receiving stations, but one has to be a bit careful as power levels increase. I recall one 40 m folded dipole excited with around 300 W of RF that got uncomfortably hot to the touch after a few minutes of CW operation.

Folded dipole construction always looks easier in books than it is in real life. The problem is that the conductors used in 300 ohm twin-lead are small, and not intended to support any weight or sustain under wind forces (which

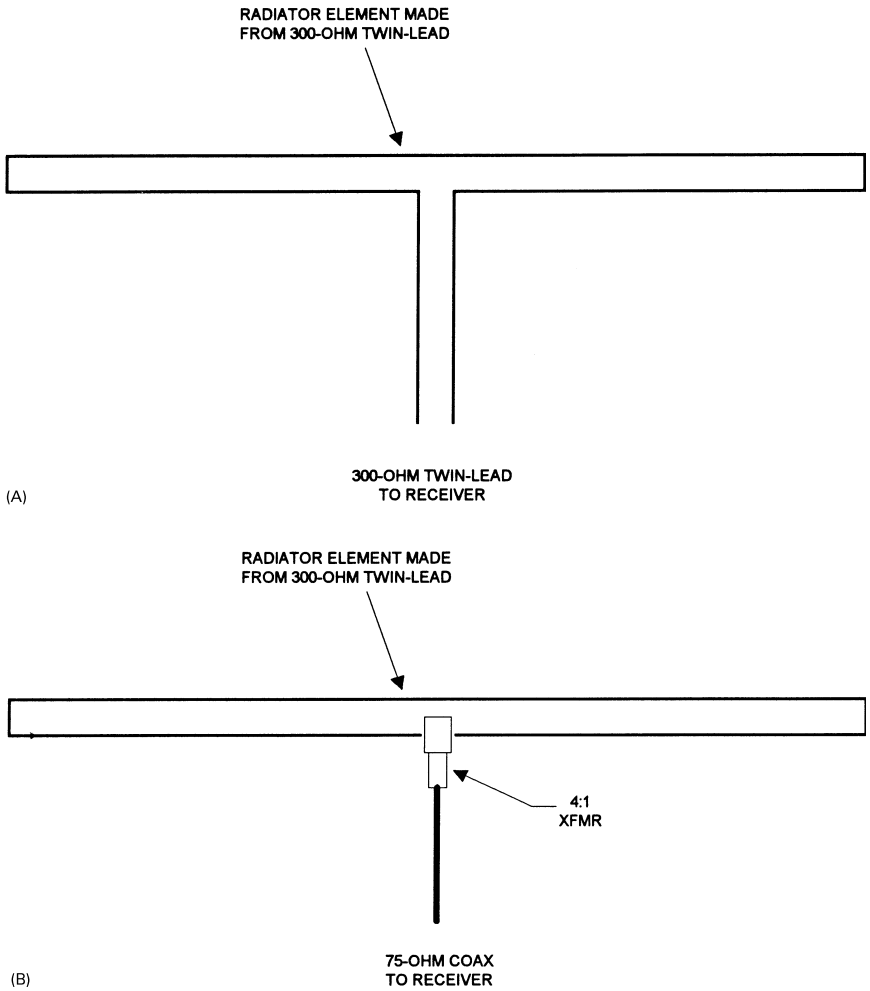


FIGURE 5.8

can be larger than most of us imagine because an antenna has a large 'sail area' in wind). As a result, folded dipoles usually break and come down either at the feedpoint or one of the two ends. The solution is to make fittings such as shown in Figure 5.9 for the center (Figure 5.9A) and ends (Figure 5.9B).

The folded dipole fittings of Figure 5.9 are made of an insulating material. I have seen similar devices made of hardwood salvaged from flooring, and then coated in polyurethene or varnish to protect against rain. Others use Plexiglass, Lexan, or other materials. I have even seen one commercial folded dipole center insulator made of nylon, but it seems to have disap-

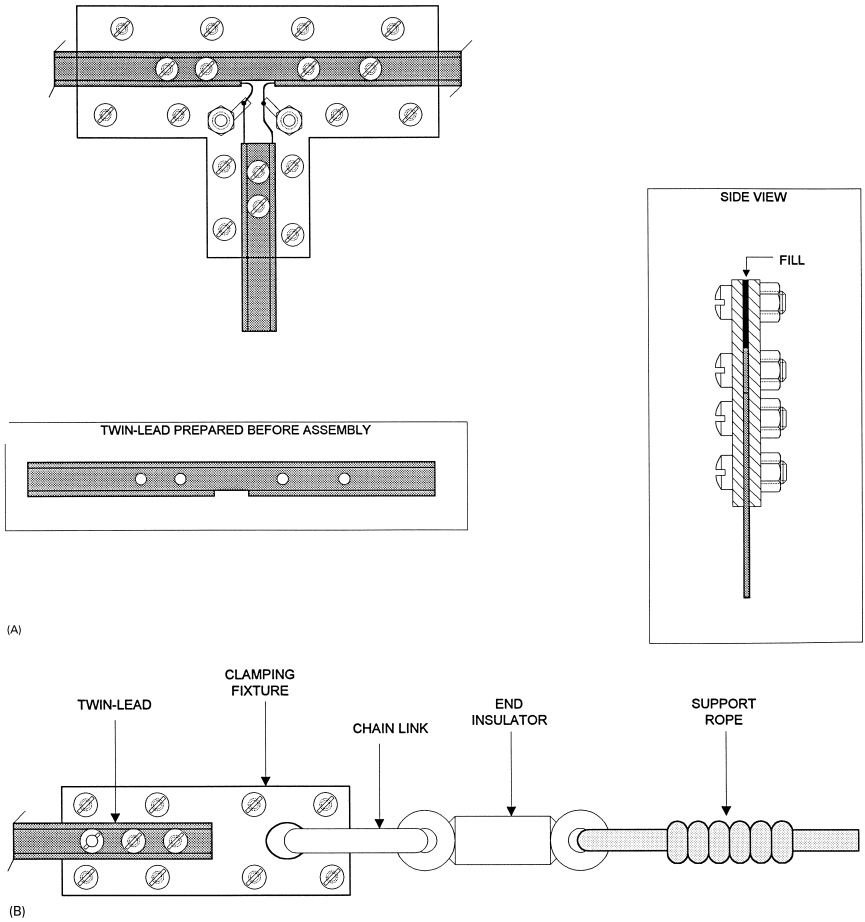


FIGURE 5.9

peared from the marketplace. In Figure 5.9A, the wire connections are soldered to solder lugs attached to brass machine screws. All other screws and nuts are made of nylon to minimize interaction with the antenna. Note in Figures 5.9A and 5.9B that holes are cut into the insulation of the twin-lead, and nylon machine screws are passed through both these holes and mating holes cut into the mounting device; matching hex nuts on the other side secure the screw (see side view inset to Figure 5.9A). This keeps the twin-lead from slipping out of the block, or at least delays the day when failure might occur.

The screw holes in the twin-lead (see inset to Figure 5.9A) can be made with an ordinary paper hole punch, although one has to squeeze a bit hard on the better grades of twin-lead.

INVERTED-'VEE' DIPOLE

The **inverted-'vee' antenna** (Figure 5.10) is a dipole that is supported in the center from a mast or other support, with the ends drooping toward the ground. As with other forms of dipole, the ends are supported using end insulators and rope. But instead of finding a convenient high spot to support the ends, the rope is tied off to a ground level support, or a stake driven into the ground if none exists. Keep in mind, however, that only the type of tent pegs that are intended for high-wind areas (i.e. the type with prongs) should be used. Otherwise, drive a long stake into the ground. The wires of this antenna have a rather large sail area, and the forces could easily pull out the ground supports over time.

The inverted-'vee' antenna is fed in the center with 75 ohm coaxial cable and a 1:1 BALUN transformer. In some cases, you will find that 52 ohm coaxial cable is a better match. The angle between the radiator elements at the top of the mast should be 90–120°. The directivity is based on the figure '8' of a dipole, but with considerable filling in of the nulls off the ends.

The length of the wire elements can be a quarter wavelength each (a half wavelength overall), or any odd integer multiple of a quarter wavelength. Because of the sloping of the elements, the actual length is of the order of 5–6% longer than for an equivalent horizontal dipole at the same frequency. The quarter-wavelength element lengths are found from

$$L_{\text{meters}} = \frac{76}{F_{\text{MHz}}} \text{ meters}$$

As with all antennas, the actual lengths required will be found experimentally using the calculated figure as a starting point.

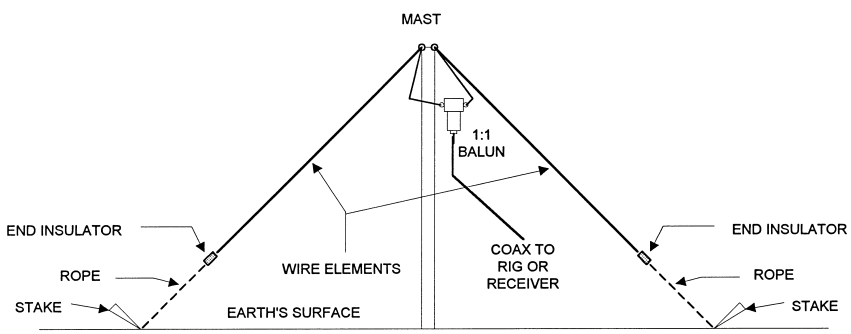


FIGURE 5.10

SLOPING DIPOLE

The **sloping dipole** (Figure 5.11) is a half wavelength dipole. Like the inverted-'vee', this dipole slants from a support (mast, tree, rooftop) to the ground. Unlike the 'vee', however, the sloping dipole has the entire half wavelength in one sloping element. The length of each quarter-wavelength element is found from the same equation as for the inverted-'vee'.

VERTICAL DIPOLE

A refinement of the sloping dipole is the vertical dipole shown in Figure 5.12. Like all verticals, the azimuthal pattern of this antenna is omnidirectional if nothing distorts it. The vertical dipole antenna works especially well in locations where there is limited space, and the desired frequencies are in the higher end of the HF spectrum. Lower-frequency vertical dipoles can be accommodated if a support of sufficient height is available. The lengths of the quarter-wavelength elements are found from $150/F_{\mathrm{MHz}}$, although this length will almost certainly need trimming to find the actual length.

The feed shown in Figure 5.12 is a direct feed with 75 ohm coaxial cable. If you wish to substitute a 1:1 BALUN, however, feel free. Indeed, it is recommended.

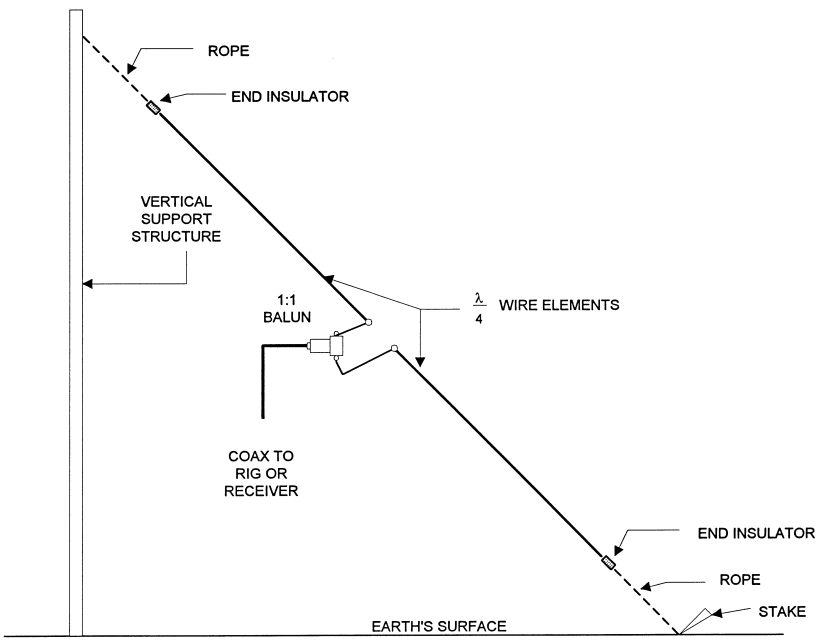


FIGURE 5.11

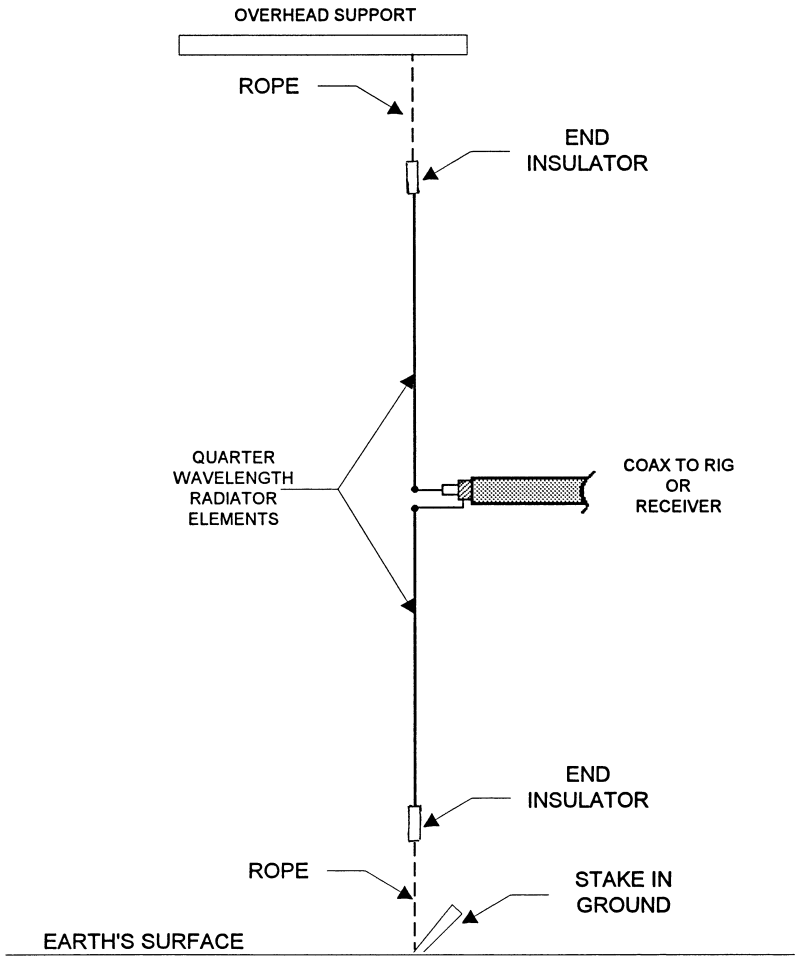


FIGURE 5.12

DELTA-FED DIPOLE

Most of the dipoles shown thus far are fed with coaxial cable. The **delta-fed half-wavelength dipole** (Figure 5.13) uses parallel feedline (either open line or 450 ohm twin-lead) and a delta match scheme. The feedline is non-resonant, and must be connected to an antenna tuning unit that has a balanced output. The lengths are found from

$$A_{\text{meters}} = \frac{142}{F_{\text{MHz}}} \text{ meters}$$

$$B_{\text{meters}} = \frac{54}{F_{\text{MHz}}} \text{ meters}$$

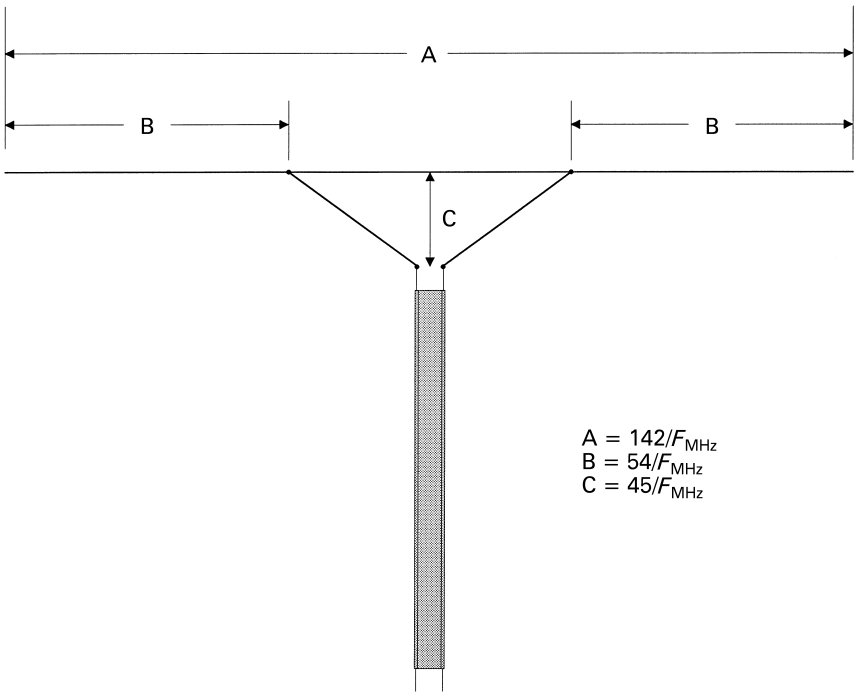


FIGURE 5.13

$$C_{\text{meters}} = \frac{45}{F_{\text{MHz}}} \text{ meters}$$

The delta-fed dipole was very popular prior to World War II, and still finds some adherents.

BOW-TIE DIPOLE

Figure 5.14 shows the 'bow-tie dipole' wide-band antenna. It is not quite like the wide-band antenna shown in Figure 5.6 because both wires in both elements are cut to the same frequency. Wide-band operation is achieved by spreading the ends of the two wires in each element approximately 11% of the total length. Length L (in meters) is found from $127/F_{\text{MHz}}$, and the spread width $W = 0.11L$. For a frequency of 4.5 MHz, therefore, the overall length of the antenna is $127/4.5 = 28.22$ m, so $W = 0.11 \times 28.22 = 3.1$ m. The bandwidth of the antenna is achieved because spreading the ends supposedly controls impedance excursions that would normally occur when the frequency departs from resonance.

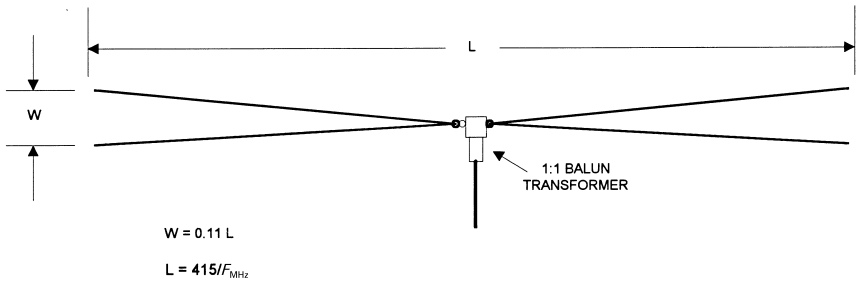


FIGURE 5.14

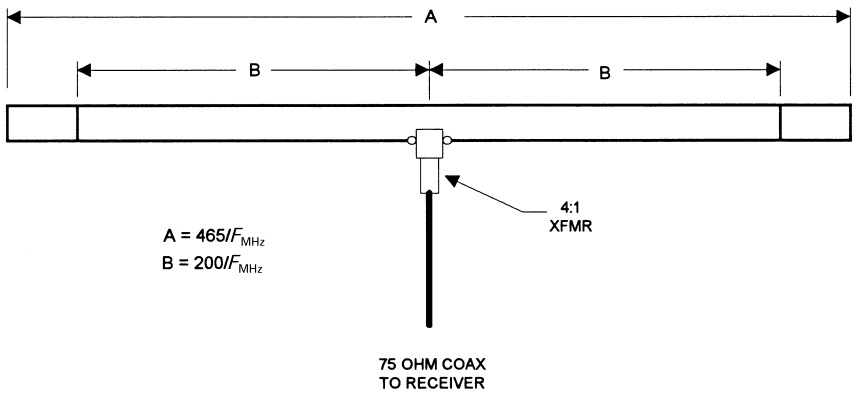


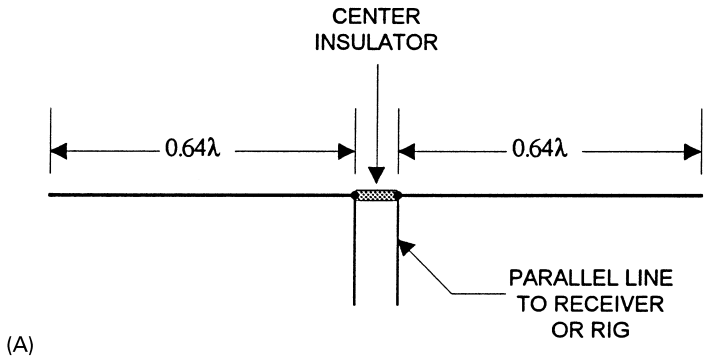
FIGURE 5.15

WIDER BANDWIDTH FOLDED DIPOLE

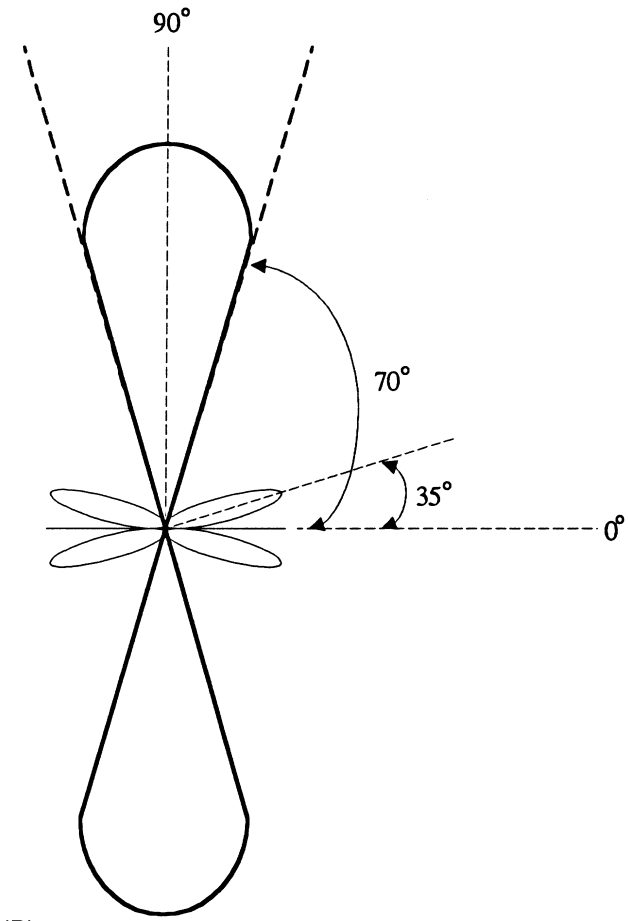
Figure 5.15 shows a wider bandwidth folded dipole antenna. This antenna is made from 300 ohm twin-lead. Like all folded dipoles, the parallel conductors are shorted at the ends, and one conductor is broken at its center to accommodate either a 300 ohm twin-lead feedline to the receiver, or a 75 ohm coaxial cable coupled to the antenna through a 4:1 BALUN transformer. In the antenna of Figure 5.15, however, there is a difference from the normal folded dipole format: shorts are placed between the two conductors at very critical distances (B) from the feedpoint. The overall length (A) of the folded dipole is given by $142/F_{\mathrm{MHz}}$, while the distance between the inner shorts (B–B) is $122/F_{\mathrm{MHz}}$, i.e. the distance from the feedpoint to each inner short is $61/F_{\mathrm{MHz}}$.

DOUBLE EXTENDED ZEPP

The double extended Zepp antenna (Figure 5.16A) is a dipole in which the overall length is 1.28λ, and each element is 0.64λ. It provides about 3 dB



(A)



(B)

FIGURE 5.16

gain over a dipole. The azimuthal pattern (Figure 5.16B) is a narrowed figure '8' with sidelobes appearing at $\pm 35^\circ$ from the line along the antenna length.

The length of each 0.64λ element is found from

$$L_{\text{meters}} = \frac{163}{F_{\text{MHz}}} \text{ meters}$$

This antenna must be fed with parallel line connected to an antenna tuning unit with a balanced output. However, if you build a one-eighth wavelength matching section from 450 ohm twin-lead, then you can reduce the impedance to 150 ohms. The length of the 450 ohm twin-lead matching section is

$$L_{\text{meters}} = \frac{31.4}{F_{\text{MHz}}} \text{ meters}$$

The impedance looking into the matching section is 150 ohms, so using 75 ohm coaxial cable results in a 2:1 VSWR. However, if a 4:1 BALUN transformer is provided, then the impedance is transformed to $150/4 = 37.5$ ohms. If 52 ohm coaxial cable is used with this scheme, then the VSWR $= 52/37.5 = 1.4{:}1$.

MULTIBAND TUNED DOUBLETS

An antenna that was quite popular before World War II with ham operators and short-wave listeners is still a strong contender today for those who want a multiband antenna, provided that the bands are harmonically related. Figure 5.17 shows the **multiband tuned doublet** antenna. This antenna consists of a nominal half-wavelength wire radiator that has a physical length (in meters) of $145/F_{\text{MHz}}$, where F_{MHz} is the *lowest* frequency (expressed in megahertz) of operation. Note this difference: many antennas are designed for the center frequency of the band of interest, while this one is designed for the lowest frequency. The antenna will perform well on harmonics of this frequency, and will perform at least somewhat on other frequencies as well if a higher VSWR can be tolerated. This situation is seen for receiver operators, but is often a limiting case for transmitter operators.

The feed of this antenna is through a quarter-wavelength matching section made of 450 ohm twin-lead transmission line. The length of the quarter-wavelength section is $73/F_{\text{MHz}}$. Although one may get away with replacing the heavier 450 ohm line with lighter 300 ohm line, it is not appropriate to replace the line with coaxial cable and a 4:1 BALUN as is done in other antennas. The reason for this is that the twin-lead forms a set of tuned feeders, so is not easily replaced with an untuned form of line.

The lengths for typical antennas of this sort are given in Table 5.1.

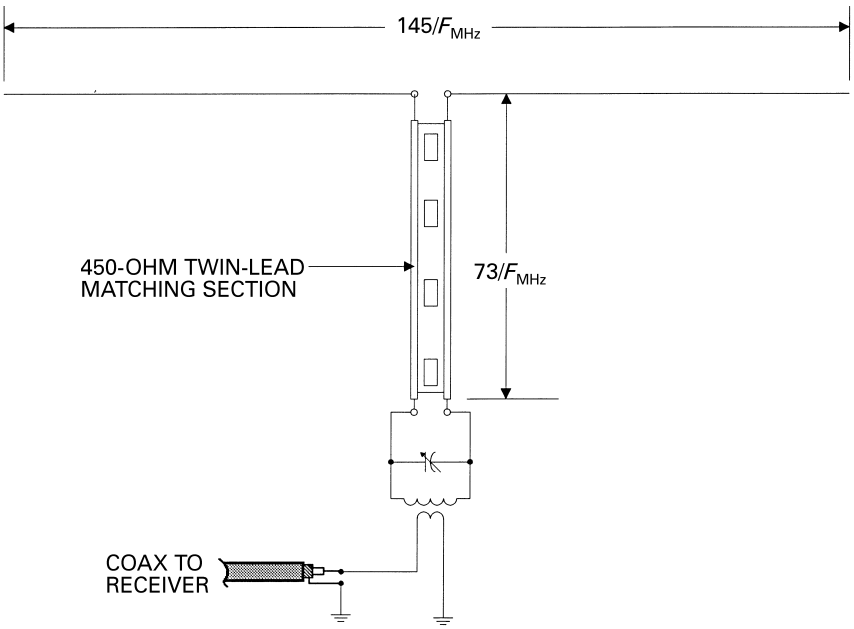


FIGURE 5.17

TABLE 5.1

Lower frequency (MHz)	Radiator element (m)	Matching section (m)
3.5	41.43	20.86
5.0	29.00	14.60
7.0	20.71	10.43

One consequence of this configuration is that a balanced antenna tuner is needed at the feed end of the matching section. The version shown here is a parallel resonant LC tank circuit that is a transformer coupled to a low-impedance link to the receiver or transmitter. The inductor and capacitor for this tuner should be designed so that they have a reactance of about 600 ohms at every frequency required. This may mean a tapped or variable inductor, or bandswitched inductor, as well as a variable capacitor. Commercial antenna tuners can also be used for this application.

THE G5RV ANTENNA

The G5RV antenna (Figure 5.18) is controversial for several reasons. One reason is the originator of the antenna. The name 'G5RV' is the call-sign of the claimed originator, Louis Varney, a British ham operator. Others claim

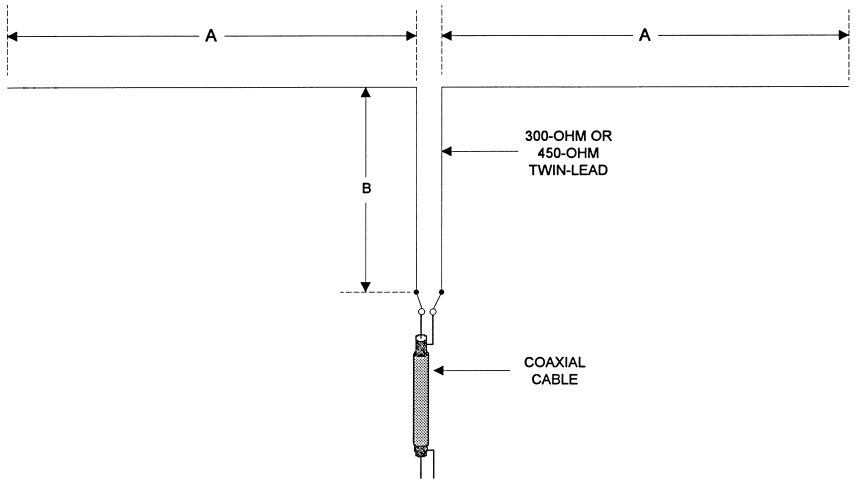


FIGURE 5.18

that the G5RV antenna is nothing but a 1930s or 1940s vintage design by Collins Radio for the US military. However, the similarities between the Collins antenna and the G5RV are, it seems to me, at best a case of 'further development' or 'co-invention,' rather than something more sinister. Because of the obvious differences between the two antennas, I prefer to continue the credit due to G5RV.

Another reason for the seeming controversy over the G5RV is perhaps the 'NIH syndrome' (not-invented-here). The G5RV is more popular in Europe than in the USA. In my own experience, the G5RV tends to be built in the US by antenna experimenters, and most of those who I have talked to are happy with the results.

Still another controversy is over whether, or how well, the G5RV works. One reviewer of one of my other antenna books (*Joe Carr's Receiving Antenna Handbook*, Universal Radio) stated that he 'wish(es) the G5RV would just go away.' The same reviewer stated that it would be better to just put up a 'dipole of the same size.' Wrong! The dipole is a single-band resonant antenna, whereas the G5RV will work on several harmonically related bands. The G5RV has two poles, but it does not exactly fit into the same category as the half-wavelength dipole.

PHYSICAL STRUCTURE OF THE G5RV

The G5RV antenna looks like a dipole, to be sure, but its length is considerably longer. Unlike many multiband antennas, the G5RV is not cut to

the lowest frequency of operation, but rather to the middle frequency. For an HF ham band antenna, one designs it for 20 m (14 MHz).

Like the dipole, the G5RV is fed in the center. Unlike the dipole, a matching section made of 450 or 300 ohm twin-lead transmission line (450 ohm is preferred) is connected between the antenna feedpoint and the 75 ohm coaxial cable. The length of each radiator element (A) is

$$A = \frac{220}{F_{\text{MHz}}} \text{ meters}$$

While the length of the matching section (B) is

$$B = \frac{146V}{F_{\text{MHz}}} \text{ meters}$$

A is the length of each radiator element, B is the length of the 450 ohm matching section, F_{MHz} is the middle frequency of operation, and V is the **velocity factor** of the twin-lead (typically 0.82 for twin-lead and 0.99 for open-wire parallel line).

In case you do not like to do arithmetic, the calculations have already been done for the HF ham bands: $A = 15.55$ meters ($2A = 31.1$ m overall), and $B = 10.37$ m for open-wire transmission line and 8.38 m for 300 or 450 ohm twin-lead. There is some argument over these figures, but they are regarded by many hams who have actually used the antenna as a good trade-off.

If you want more technical details on the G5RV, Louis Varney, the inventor, has written about the antenna in a number of publications. Varney gives the basis for operation of the G5RV antenna at 3.5, 7, 10, 14, 18, 21, 24, and 28 MHz. The VSWR on each band is a bit different, and Varney recommends the use of a transmatch or similar coaxial-to-coaxial antenna tuning unit between the transmitter and the input to the coaxial cable transmission line.

There is a possibility of an unbalanced line condition existing that causes some radiation from the feedline – and for transmitters that can cause television interference (TVI) and other forms of unpleasantness. The solution to this problem is to wind the coaxial cable into an in-line choke at the point where the coaxial cable connects to the twin-lead or parallel line matching section. This is done by winding the coaxial cable into a 15 cm diameter coil of 10 turns right at the feedpoint. The coiled coaxial cable can be secured to the center insulator by tape, string, or some other mechanism.

BALUNS ON RECEIVER ANTENNAS?

Above I mentioned a reviewer who did not like the G5RV antenna. The same reviewer stated that he did not believe that 1:1 BALUN transformers are useful on half-wavelength dipole receiving antennas. My calm, professional, well-considered response is: *rubbish*! The purpose of using a 1:1 BALUN (which after all provides no impedance transformation) is twofold. First, as in the transmit case, the BALUN prevents radiation from the transmission line. Perhaps the reviewer was thinking of TVI as the reason for avoiding feedline radiation. But, as ample test chamber evidence shows, the radiation from the feedline tends to distort the figure '8' azimuthal pattern. When a BALUN is used, the currents are balanced, and the radiation pattern is restored. And guess what? Antennas are reciprocal in nature – they work the same on receive as on transmit.

The second reason is that the receiver antenna feedline may pick up strong signals from powerful local stations. Other hams and AM broadcast band stations are particular problems. Any large signal at the input may challenge even the best receiver front-end, but if the receiver design is in any way mediocre in the dynamic range department (and many are!), then the signals picked up on the dipole transmission line shield can overload the front-end of the receiver. Using a 1:1 BALUN transformer between the antenna feedline and the radiator elements balances out the currents and seriously reduces the amount of signal seen by the receiver input.

LINEAR LOADED WIDE-BAND DOUBLET

Figure 5.19 shows an antenna that superficially resembles both the folded dipole and the G5RV, but is actually a wide-band antenna with a linear loading section between the radiators. The overall length of the main radiator elements is $146/F_{\mathrm{MHz}}$ meters, while the linear loading element is $133/F_{\mathrm{MHz}}$ meters long. The radiator elements are separated about 30.5 cm, with the linear loading element placed in the middle, 15 cm from each element. The linear loading element is placed horizontally about the mid-point of the antenna. The spreaders can be ceramic, or made of any insulating material. In one variant that I built, ordinary wooden dowels (1 cm diameter) were used, although I am not too sure it will weather well unless treated with some kind of coating.

This antenna is fed through 'any' length of 450 ohm twin-lead, but practical advice from several people who have actually built this antenna suggest that 'any' should be read as 'one-fifth to one-half the length of the radiator elements.' A 9:1 BALUN transformer is connected to the feedpoint, and coaxial cable routed to the receiver.

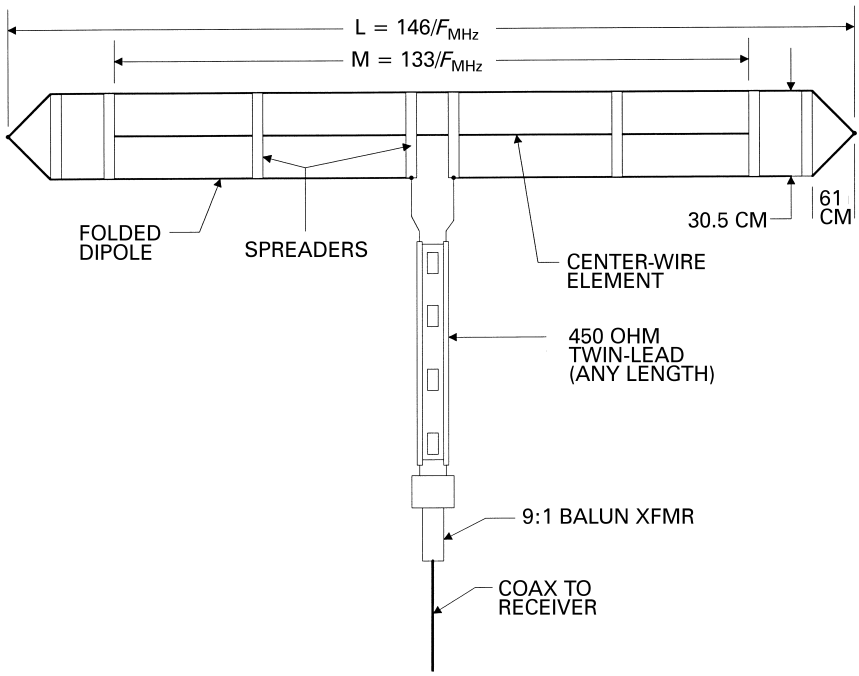


FIGURE 5.19

One practical problem with this antenna is that it tends to flop over because it is top heavy. Two solutions present themselves. First, do not tie off the ends in a single rope, but rather attach two ropes to the supporting structure at each end. Second, tie rope or fishing line from the bottom radiator at each end to a supporting structure below, or to stakes in the ground. One fellow used fishing weights for this purpose. The weights were attached to the bottom radiator element at the far ends.

THE AUSSIE BROAD-BAND HF DOUBLET FOR RECEIVERS

One of the problems with most of the antennas in this chapter is that they work on a single resonant frequency, and a narrow range of frequencies around it. Some of them will work over relatively wide bandwidths, but the coverage is nonetheless limited with respect to the entire HF spectrum. The Aussie doublet shown in Figure 5.20 is a receive antenna with a VSWR less than 2.5:1 over the entire 3–30 MHz range of the HF band.

This antenna is rather large, being 40 m long and 1.8 m wide, so requires some space to erect. Also, the two conductors of each segment are separated by insulating spacers. One approach to making these very long (1.8 m) spacers is to use PVC plumbing pipe or small size lumber. Even then, the

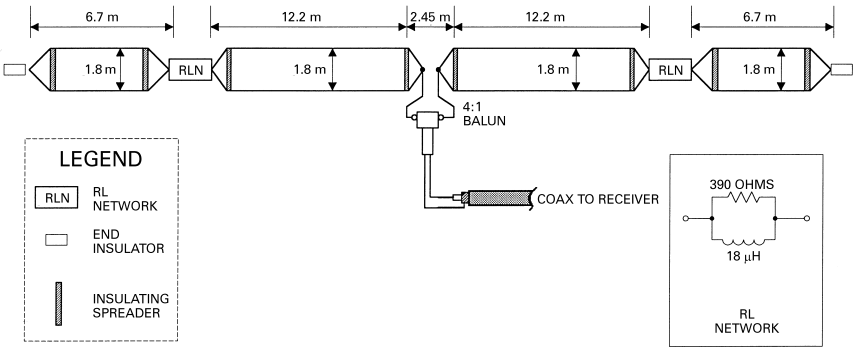


FIGURE 5.20

antenna will tend to flop over and lay flat unless rope moorings to the ground are provided along the bottom conductor.

Part of the broad-banding effect is provided by the two-wire design and their spacing, but much of it is also provided by a pair of RL networks ('RLN' in Figure 5.20). Each network is a 390 ohm resistor in parallel with an 18 μH inductor. It is these components that limit the use of this antenna for transmitting, although moderate powers can be accommodated if a suitable non-inductive, high-power resistor and a suitably designed inductor are provided.

THE CAPACITOR-TUNED WIDE-BAND DIPOLE

Dipole antennas are resonant and so will operate effectively over a relatively narrow band, as well as the third harmonic of that band. It is possible to broaden the response of the antenna by inserting a variable capacitor at the feedpoint (Figure 5.21). This capacitor should have a maximum value of 500 pF, although 730 and 1100 pF have been successfully used. The latter two values represent the values obtained when two or three sections of a 365 pF broadcast variable capacitor are used (receive only). For transmitting, a wide air gap capacitor ('transmitting variables') or vacuum capacitor should be used.

The best solution for using this antenna is to mount the capacitor at the feedpoint inside of a weatherproof housing, and then use a low-voltage, low-RPM direct-current (DC) motor to drive it. Voltage to the motor can be supplied through a separate wire, using the coaxial cable shield as the DC return line. In another scheme, it is also possible to use the coaxial cable center conductor to carry the DC (with the shield acting as the return), but it is necessary to separate the motor and signal line with an RF choke at both

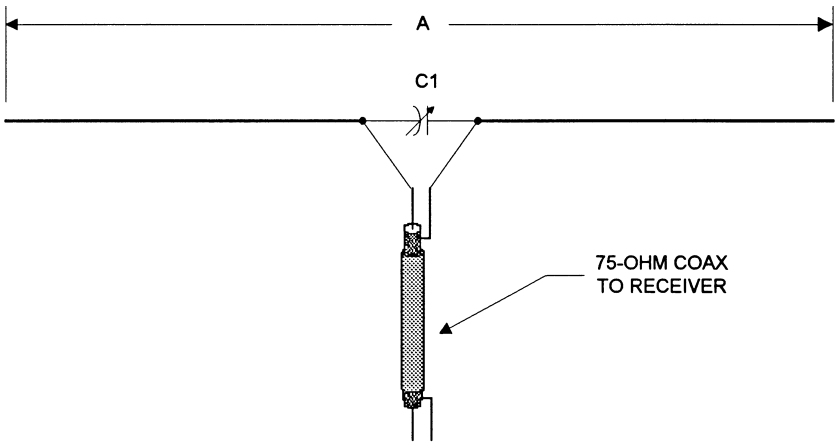


FIGURE 5.21

ends of the transmission line. IT IS VERY IMPORTANT TO USE LOW-VOLTAGE DC FOR THIS JOB. DO NOT USE AC POWER MAINS VOLTAGE!

THE INDUCTOR-LOADED SHORTENED DIPOLE

At low frequencies, dipole antennas can take up a large amount of space. At 75/80 m ham bands, for example, a half-wavelength dipole is 41 m long; at 5 MHz, the length is still 28.5 m. This fact of physics can ruin the plans of many radio enthusiasts because of space limitations. There is, however, a method for making a dipole one-half the normal size. Figure 5.22 shows a dipole that is one-half the size of the normal half-wavelength dipole, even though it simulates the electrical half wavelength.

An antenna that is too short for its operating frequency offers a capactive reactance component to the impedance at the feedpoint. In order to tune out this impedance the opposite form of reactance, inductive reactance, is placed in series with each radiator element. Although the inductors can be placed anywhere along the line, and the overall length can be any fraction of a 'full-

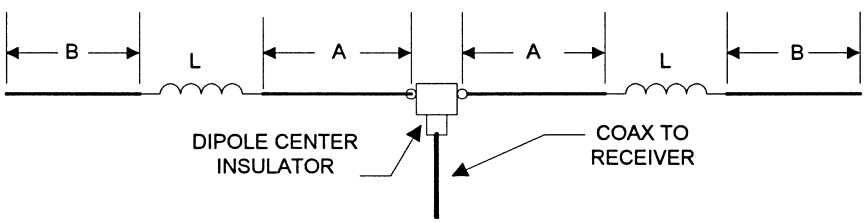


FIGURE 5.22

size' half-wavelength, the design gets a bit messy. A compromise is found by limiting our selections: (1) the overall length is one-half the normal length; and (2) the inductors are placed at the center of each element. If these rules are followed, then we have a shot at making this antenna with only a small amount of discomfort in the math zone. In this limited case, the reactance of each coil is 950 ohms in the middle of the band of interest. The procedure is as follows:

(a) Calculate the overall length of the elements: $L_{\text{meters}} = 37.5/F_{\text{MHz}}$. This is the overall length of the half-size antenna.
(b) Divide the number obtained in step (a) by 2. This is the length of each segment of wire (A or B). In our limited case, the assumption is that $A = B$, but in some other cases this assumption might not be true.
(c) Calculate the inductance required to produce a 950 ohm reactance at the design frequency:

$$L_{\mu\text{H}} = \frac{X_{\text{L}} \times 10^6}{2\pi F_{\text{MHz}}}$$

Antennas do not give up convenience for free, however. The price of this antenna design is narrower bandwidth.

MORE MULTIBAND DIPOLES

The shortened inductively loaded dipole of Figure 5.22 can be used on two bands by increasing the reactance of the coil to 5500 ohms at the middle frequency of the highest band. For example, you can make a 75/80 m plus 40 m dipole. The inductor in this case serves not only to foreshorten the antenna's overall length, but also acts as an RF choke to the higher-frequency band. In that case, the sections labelled A in Figure 5.22 are for the higher-frequency band, while $A + B$ is for the lower-frequency band. Approximations of the lengths are found from

$$A = \frac{74}{F_{\text{HI(MHz)}}} \text{ meters}$$

$$B = \frac{4.64}{F_{\text{LO(MHz)}}} \text{ meters}$$

In this antenna, the sections A act like a regular half-wavelength dipole at the higher frequency, while the sections $A + B$ act like a shortened inductively loaded dipole at the lower frequency. You are well advised to leave extra space on the B-sections for adjustments. This antenna tends to be narrower in bandwidth on the lower band, so may require some fiddling with lengths to achieve the desired resonance.

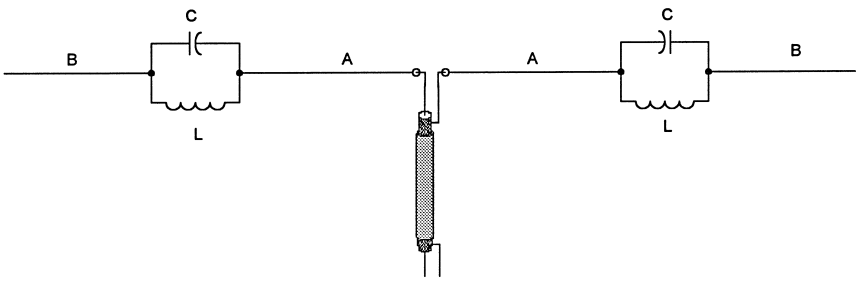


FIGURE 5.23

Another approach to multibanding is shown in Figure 5.23. This antenna is called a **trap dipole** because it uses a pair of parallel resonant LC networks to trap the higher frequency. A parallel resonant network presents a high impedance to signals at its resonant frequency, but a low impedance to frequencies removed from resonance. Like the antenna above, the sections marked A are a half-wavelength dipole at the higher frequency, while the $A + B$ sections form a near-half wavelength at the lower frequency. These lengths are approximated by the normal half-wavelength equations, although some adjustment is in order due to the effects of the trap. The traps consist of an inductor (L) and capacitor (C) in parallel. At resonance, $X_L = X_C = 370$ ohms.

MULTIBAND RESONANT ANTENNAS: MULTIBAND TUNED DOUBLET

An antenna that was quite popular before World War II is still a strong contender today for those who want a multiband antenna, provided that the bands are harmonically related. Figure 5.24 shows the multiband tuned doublet antenna. It consists of a nominal half wavelength wire radiator. The antenna will perform well on harmonics of this frequency, and will perform at least somewhat on other frequencies as well if a higher VSWR can be tolerated.

The feed of this antenna is through a matching section of 450 ohm twin-lead transmission line. Although one may get away with replacing the heavier 450 ohm line with lighter 300 ohm line, it's not appropriate to replace the line with coaxial cable and a 4:1 BALUN as is done in other antennas. The reason for this is that the twin-lead forms a set of tuned feeders, so is not easily replaced with an untuned form of line. One consequence of this configuration is that a balanced antenna tuner is needed at the feed end of the matching section. The version shown here is a parallel

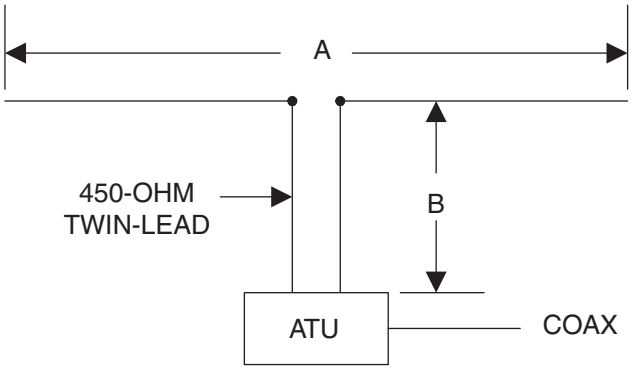


FIGURE 5.24

resonant L-C tank circuit that is transformer coupled to a low impedance link to the receiver.

The antenna is a half wavelength long at the lowest frequency of operation, and the twin-lead is a quarter wavelength long, or an odd integer multiple of quarter wavelengths. The lengths are:

$$A = \frac{144}{F_{\text{MHz}}}$$

$$B = \frac{73}{F_{\text{MHz}}}$$

This antenna depends on an antenna tuning unit (ATU) to be effective, unless the transmitter has a balanced output that can handle the high impedance. The ATU needs to be the type that has a balanced output, rather than the coaxial cable 'line flattener' variety.

THE TILTED, CENTER-FED TERMINATED, FOLDED DIPOLE

Figure 5.25 and the following figures show the *tilted, center-fed terminated, folded dipole* (TCFTFD) antenna, which is a special case of a loop antenna and a folded dipole antenna. The inventor, Navy captain G.L. Countryman (W3HH), once called it a 'squashed rhombic' antenna. The antenna is a widely spread folded dipole, and is shorter than a conventional folded dipole. It must be mounted as a sloper, with an angle from its upper vertical support of 20 to 40°.

The feeding of the antenna is conventional, with a feedpoint impedance close to 300 Ω. A 75 Ω coaxial cable is connected to the bottom half of the antenna through a BALUN transformer that has a 4:1 impedance ratio.

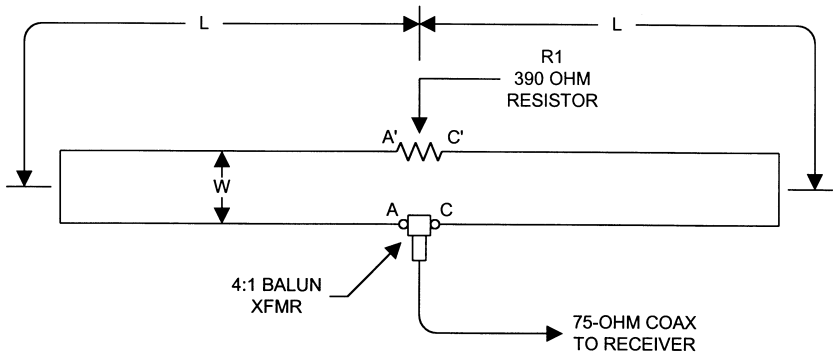


FIGURE 5.25

At the top side of the antenna, the 'feedpoint' is occupied with a termination resistor of 370 to 430 Ω (390 Ω, 1 or 2 watts, makes a good compromise for receiving antennas).

The spread (W) of the antenna wire elements is found from:

$$W = \frac{2.99}{F_{\text{MHz}}}$$

where W is the width of the antenna in meters (m) and F_{MHz} is the frequency in megahertz.

The spreaders are preferably ceramic, strong plastic, or thick-walled PVC pipe. The spreaders can be made of wood (1 × 2 stock or 1-inch dowels) for receive antennas if the wood is properly varnished against the weather.

The overall length of the antenna is calculated a little differently from most antennas. We need to calculate the lengths from the *feedpoint to the middle of the spreaders*, which is also the length from the middle of the spreaders and the terminating resistor. These lengths are found from:

$$L = \frac{54.3}{F_{\text{MHz}}}$$

where L is the length in meters (m) and F_{MHz} is the frequency in megahertz.

Four sections of wire, each with a length defined by the equation above, are needed to make this antenna.

The height of the upper antenna support is determined by trigonometry from the length of the antenna from end-to-end (not the length calculated for D, but approximately twice that length), and the angle. For example, at 7 MHz the lengths are 7.76 meters, and the spreaders are 0.47 cm. Thus, the overall physical length, counting the two element lengths and half of both spreader lengths, is [2 × 7.76 meters – (2 × 0.5)] meters, or 16.52 meters. If the angle of mounting is 30°, then the antenna forms the hypotenuse of a 60/30 right angled triangle. If we allow two meters for the lower support, then the upper support is:

$$\text{height} = 2 + L\cos Q$$
$$= 2 + (16.52\cos 60) = 10.26 \text{ meters}$$

This antenna has a low angle of radiation, and at a tilt angle of 30° (considered ideal) it is nearly omnidirectional.

The termination resistor can be mounted on a small piece of plastic, or alternatively, as shown in Figure 5.26, it can be stretched across the end insulator in the manner of the inductors in the previous section. Use a 390 Ω, 2 watt resistor for this application.

Figure 5.27 shows the installation of the antenna. The upper support and lower support are made of wood, or some other insulating material. There are ropes tying the ends of the antenna to the supports. Detail of the spreaders and end insulators are shown in Figures 5.28 and 5.29, respectively.

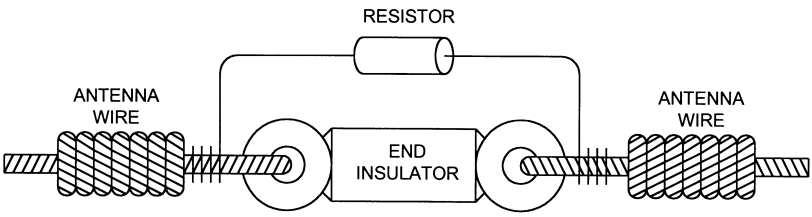


FIGURE 5.26

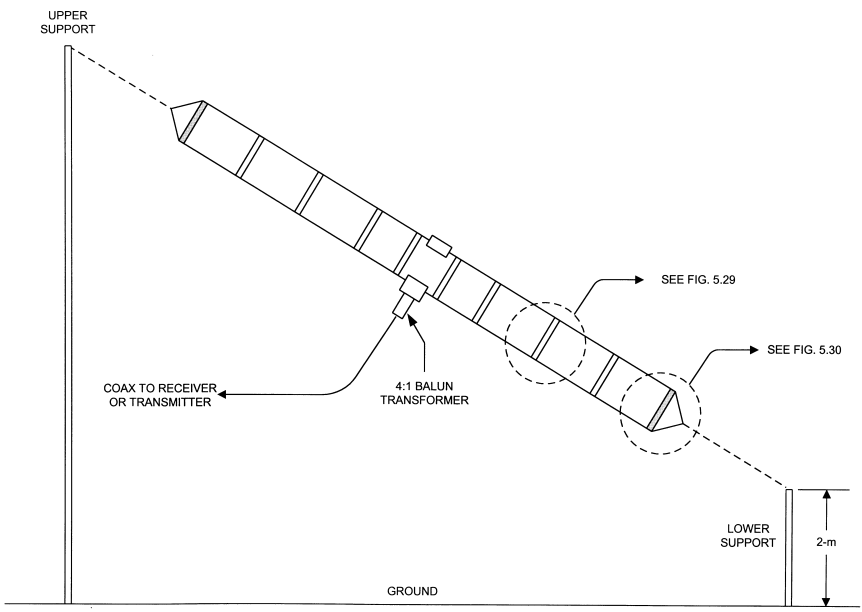


FIGURE 5.27

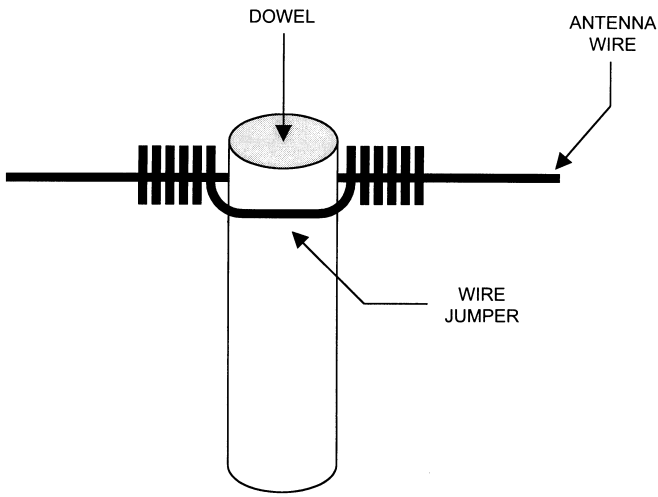


FIGURE 5.28

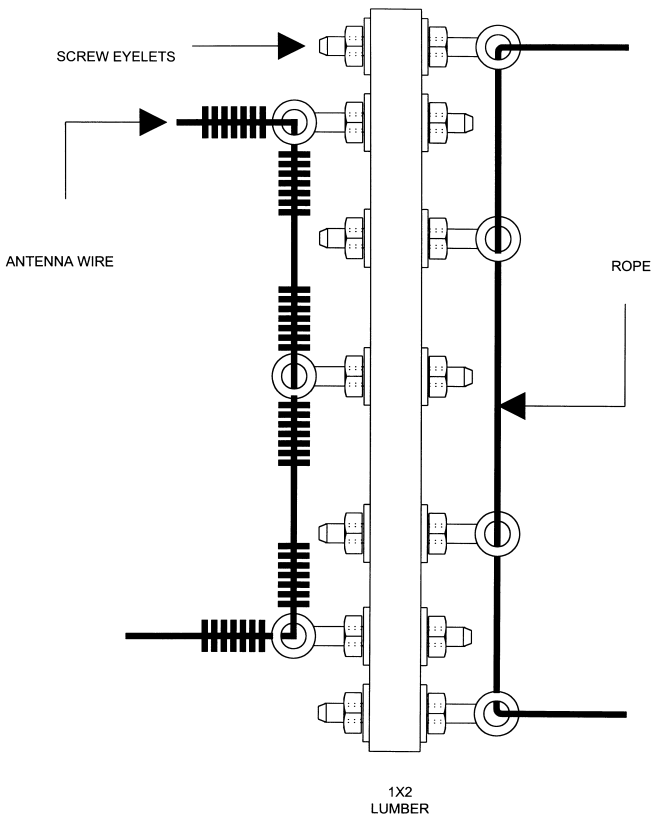


FIGURE 5.29

CHAPTER 6

Limited space antennas

Sometimes at night, when things are well and all is right in the universe, I dream. I dream of owning (or at least renting from a radio enthusiast landlord) about 500 hectares of flat land good for farming. Not that I intend to do any farming (unless 'antenna farm' counts), but good farmland is moist and composed of enough conductive minerals to make a very good radio frequency (RF) ground for antennas!

Dreams are fun to indulge, but on awakening one finds the sometimes harsh nightmare of antenna construction on the limited estates that most of us can afford. And I have lived in some tight spots. In my novice days, living in my parents' home, the lot was respectable by American suburban standards, but was criss-crossed with alternating-current (AC) power lines to both our house and the houses of two neighbors. That problem sharply limited the area open to safe antenna erection. The student boarding house where I lived in college had a sympathetic landlady (who in any event was too poor of sight to notice an antenna short of a beam and tower) was on a shallow, narrow city lot with few possibilities below 20 m (and even that band strained the antenna constructor's imagination). After being married, I lived in an apartment and then a small house of my own for a number of years – all of which were limited spaces. Even with my present home, which seems a mansion compared to our first house, my ability to put up full size antennas is a testament to my wife's tolerant attitude. So what to do?

Fortunately, there are some things that can be done to accommodate the limited space situation. Nearly all of these schemes use some form of com-

pensation antenna, and as a result do not work as well as the full-sized equivalent. The TANSTAAFL principle reigns supreme: '*there ain't no such thing as a free lunch*.' Whenever one attempts to reduce the size of an antenna for any given frequency, the result is that something else falls down a bit: gain, directivity, impedance, efficiency, bandwidth, or all of the above. While these factors may lead one to a depressed sense of gloom and doom, that is not the best attitude. The correct way to view the situation is not comparing the limited space antenna to a full size antenna, or the best in class, but rather to whether or not it allows you to operate at all. After all, a 10 m long, end-fed inductor loaded Marconi draped out of an upper-story window may not work as well as a 75 m dipole, 25 m in the air, but it works well enough to make contacts.

SOME OBVIOUS SOLUTIONS

Not all limited space antenna choices are bad. For example, one could restrict operations to the upper bands, i.e. those above 21 MHz and into the very high-frequency (VHF) region. I know one fellow who was a 300+ country DXCC award holder who operated only on the 15 m ham band because of space limitations (he lived on a small $\frac{1}{12}$ hectare lot). He was able to erect a 15 m two-element quad beam that worked quite well. Another fellow earned his DXCC as a newly licensed amateur (when operating skills are presumedly undeveloped as yet) with a 50 W kit-built transmitter and a 10–15–20 m trap vertical mounted on the roof of a four-story apartment building. It can be done.

The vertical antenna is a decent solution for many people. A full-sized 40 m (7–7.3 MHz) vertical is only 10 m high, and that is right at the legal limit or 10.7 m in my county for an installation that does not need mechanical inspection by the local government. Adding a switchable loading coil at the base of that antenna would also allow 75/80 m operation, especially if an antenna tuning unit is provided.

Modern trap verticals are also a good bet. Several commercially made verticals allow operation over the 40 m band through 10 m and are only 6.5 m tall when mounted on the ground. A good set of radials makes this antenna reasonably efficient. The only drawback is that the omnidirectional pattern makes interference rather constant. The half-wavelength bottom-fed ham-band verticals currently on the market are quite decent DX antennas.

Another obvious solution is to use any of several inductor loaded antennas such as the loops and shortened dipoles shown elsewhere in this book. Some of these antennas can be quite useful. Also, I once operated with a pair of commercial coil-loaded mobile antennas mounted back to back to form a kind of dipole. Did it work well? No! But it worked well enough for me to rack up a lot of DX using an old Heath HW-101 transceiver.

I have also used commercial mobile antennas as fixed verticals. In one case, right after we moved into our first home, I mounted a Hustler mobile antenna for 20 m on the window sill of the upstairs bedroom that served as my office (until the kids started coming along). A pair of radials out of the window, sloping to a fence, completed the 'ground' side of the antenna. It worked rather well, actually. Later on, I added 15 and 40 m coils using one of those attachments that allow three coils to coexist on the same base mast. It worked very well at 15 m and 20 m, and passably well at 40 m.

SOME OTHER SOLUTIONS

Figure 6.1 shows Marconi (Figure 6.1A) and Hertzian (Figure 6.1B) antennas for use in limited space situations. In the Marconi version, the antenna wire is mounted to lay flat on the roof, and then (if possible) to an attachment point on a support such as a mast or convenient tree. The Hertzian version is a dipole, preferably a half wavelength (if you want to use coaxial feed), positioned such that the feedpoint is in the middle of the roof line.

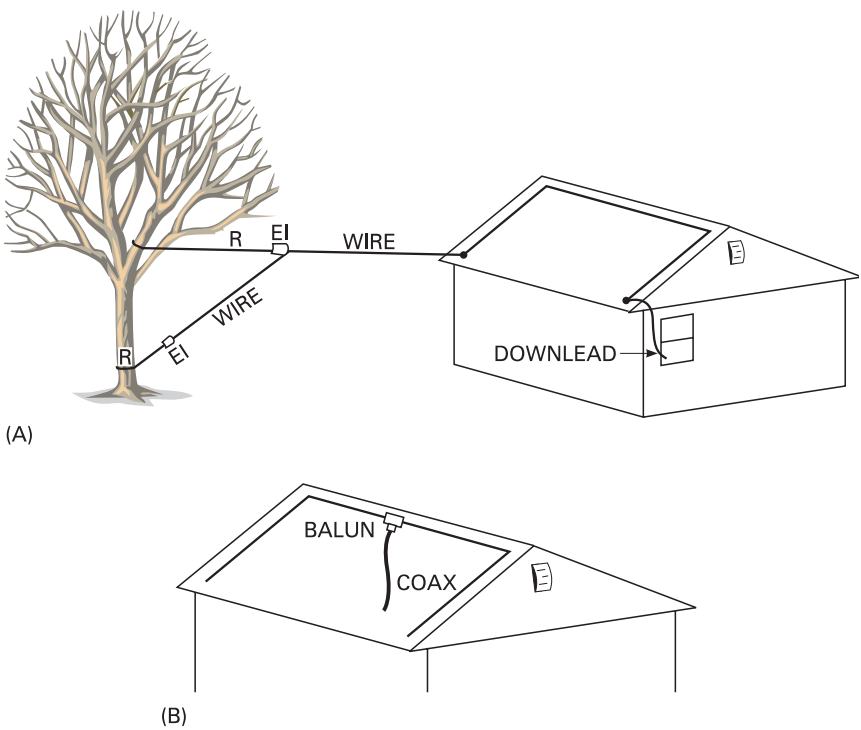


FIGURE 6.1

The idea in building these antennas is to put as much wire in the same direction as possible, but do not worry too much if the goal is not met. The real goal is to get on the air or use your receiver, depending on your interests.

Another solution is the loop antenna shown in Figure 6.2. The best loop is one that is a full wavelength on the lowest frequency of operation, but any convenient loop can be made to work at least part of the way. This is especially true if tuned parallel feeders and a decent antenna tuning unit (ATU) are used.

All of these antennas can be used either on the outside of the roof, or inside the attic. In neither case, especially the latter, would you want to try it on a roof that has a copper or aluminum base. That approach is not used on residential properties much anymore, but if you have an older dwelling the roof may well be copper covered.

Mounting the wire can be a bit of a problem, especially when the antenna is on the outside of the roof. What seems to be the best way is to use nail-in or screw-in stand-off insulators attached to the wooden roof. This works well only when you can seal it against rain water seeping in. Water will wick along the threads, and rot the wood – leading to leaks and expensive repairs. A better approach would be to design a wooden fixture attached to the soffits or overhang in a way that prevents extensive damage. Inside the attic, however, one can easily use standoffs, but they must be attached to a roof rafter (Figure 6.3), rather than the roof covering. A screw thread that penetrates to the outside surface will allow as much water damage (or more) than a thread going the other way.

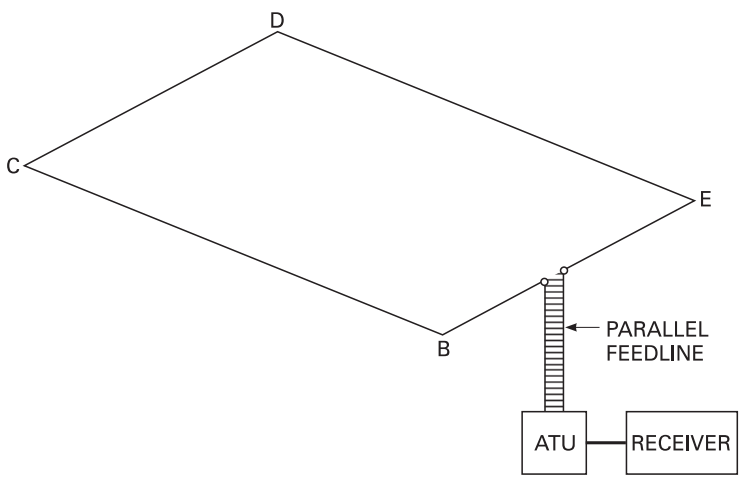


FIGURE 6.2

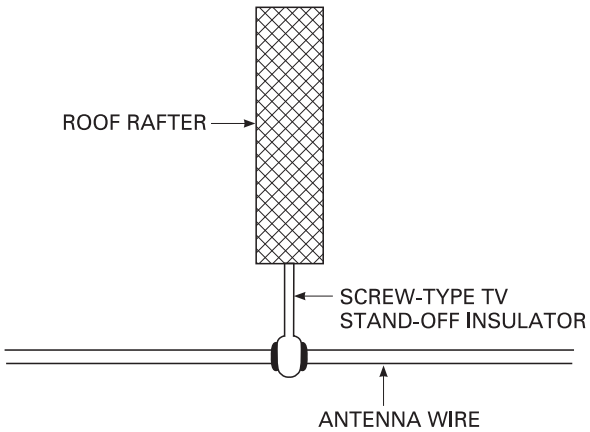


FIGURE 6.3

Marconi and Hertzian antennas may not be as good a choice as the loop because they tend to have high voltages at the ends. Use of even moderate RF power levels could cause an arcing situation that represents a fire hazard. As a result, even with the loop, if the antenna is used with a transmitter the power level should be limited.

These ideas work especially well for people with restrictive local ordinances that make decent antennas unworkable, or nasty neighbors, or restrictive covenants in their deed (and a militant homeowners association board that enforces them!). Some situations allow an antenna such as that shown in Figure 6.1A or Figure 6.1B as long as it is not visible from the street.

Zig-Zag Dipoles

The antenna purist would probably roast me over a slow fire like a marshmellow for suggesting some approaches to the limited space problem. Figure 6.4 is especially suited to drawing the ire of the pure. But it also works passably well for people with limited space. This antenna uses a zig-zag path for the dipole elements. Try to put as much wire in one direction as possible, and wherever possible make the pattern symmetrical. Although the angles shown in Figure 6.4 are acute, any angle can be used. Of course, the closer the pattern is to the regular straight dipole installation, the better it will work.

The antenna shown in Figure 6.4 is seen from the side, i.e. a ground view. The zigging and zagging can be in any direction in three-dimensional space,

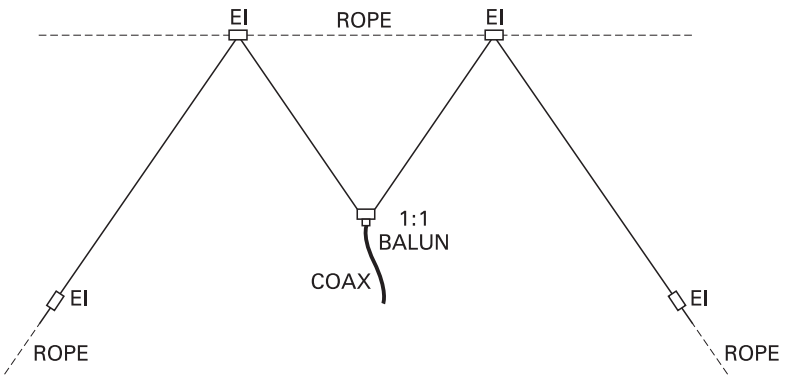


FIGURE 6.4

however. Some combinations will not work well, but others will work well enough to get you on the air.

You will find that the pattern suffers when this is done, but again the goal is to get you on the air – not to produce a perfect antenna. Also, expect to use a 'line flattening' antenna tuner to wash away the sins of the voltage standing wave ratio (VSWR).

Linearly Loaded 'Tee' Antenna

The traditional method for loading an antenna that is too short is to put a coil in series with one or all elements. In the 'tee' antenna in Figure 6.5, however, a different approach is used: **linearly loaded elements**. This antenna has been popular with low-band operators and receiver users who lack the tremendous space required for proper antennas at those frequencies. The overall length of the element (A) is usually around $\lambda/3$, and the spacing between the three segments (B) is about 25 cm (a little more or a little less can be accommodated by the design). The vertical portion of the antenna (C) is a quarter-wavelength single wire, and is fed at the bottom by parallel tuned transmission line.

The critical lengths are

$$A = \frac{50}{F_{\text{MHz}}} \text{ meters}$$

$$B = 25\,\text{cm}$$

$$C = \frac{75}{F_{\text{MHz}}} \text{ meters}$$

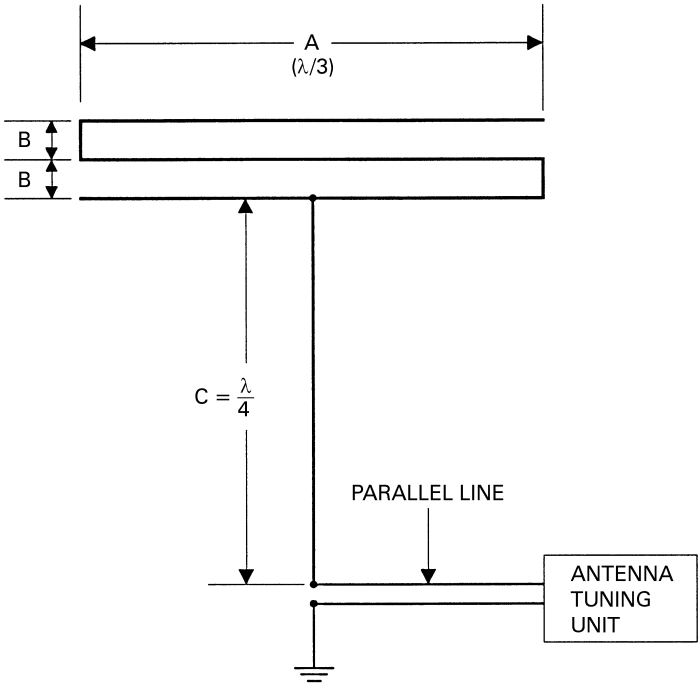


FIGURE 6.5

The Right-Angle Marconi

Figure 6.6 shows a simple quarter-wavelength Marconi antenna that can be used with 52 or 75 ohm coaxial cable. The antenna consists of two wire sections, each $\lambda/8$ long, erected at right angles to each other. The right-angle Marconi (sometimes laughingly called 'half-Hertzian') produces both a vertically and horizontally polarized signal. This method of construction reduces the linear property needed for the antenna by half, so is more accessible to a larger number of people.

Shortened Hertzian Radiators

The zig-zag dipole shown earlier is only one possibility for reducing the size of a Hertzian antenna. The antenna of Figure 6.7 is a species of zig-zag doublet, but is a little less random in its construction than the previous antenna of this sort. The overall length is $\lambda/2$, but the two $\lambda/4$ elements are bent into right-angle sections of $\lambda/12$ (vertical) and $\lambda/6$ (horizontal). The antenna is fed with open-wire parallel line operated as tuned feeders.

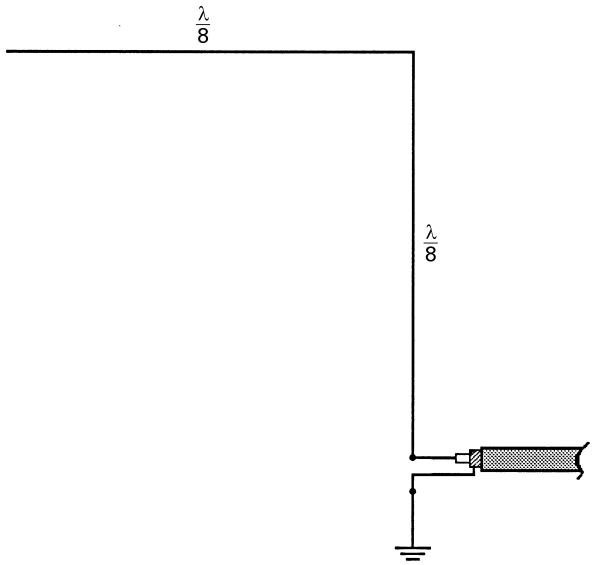


FIGURE 6.6

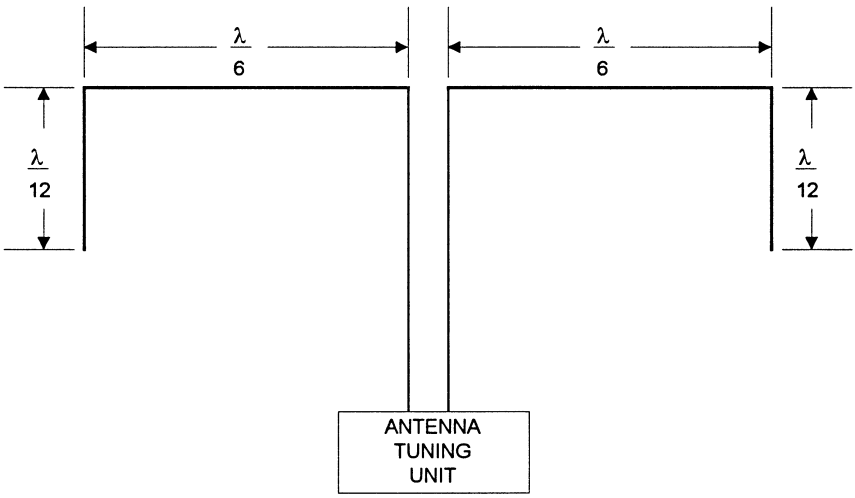


FIGURE 6.7

The antenna in Figure 6.8 looks like a folded dipole, and on one hand that is exactly what it is. The antenna length is cut for $\lambda/4$ at the lowest frequency of operation. At the second harmonic of this low band the antenna acts like a folded dipole, while at the lower frequency it is a species of 'tee' antenna. It is fed with open-wire tuned feeders. If you want to

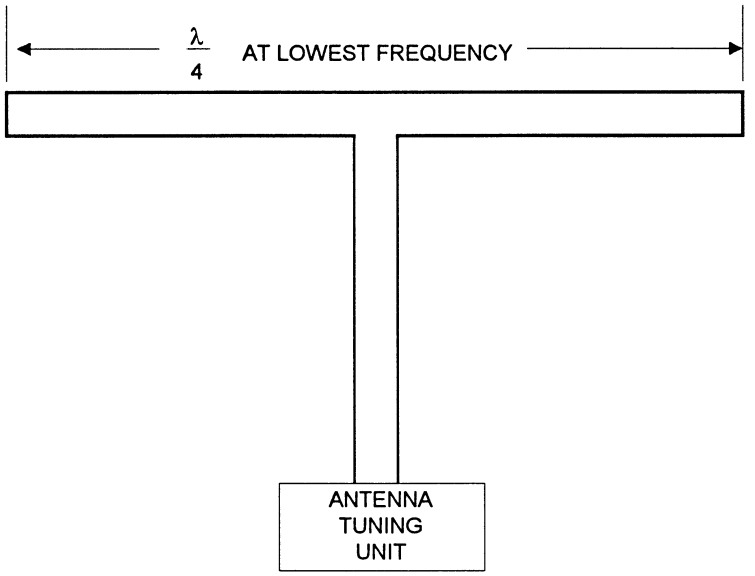


FIGURE 6.8

operate it only one band (the lower band), and wish to use coaxial cable, then a quarter-wavelength matching stub can be provided, and the antenna driven with a 4:1 BALUN transformer.

The doublet antenna in Figure 6.9 is a bit longer than $5\lambda/8$, and is fed by coaxial cable through a shorted matching stub and a 4:1 BALUN transformer. The overall length (B) of the radiator element is

$$B = \frac{102}{F_{\text{MHz}}} \text{ meters}$$

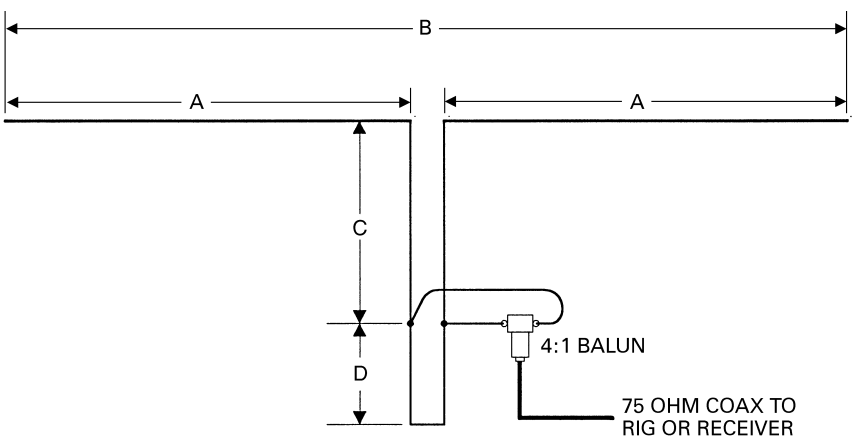


FIGURE 6.9

The matching stub consists of a length of parallel open-wire transmission line, divided into two segments either side of the feedpoint. The dimensions of the two segments are

$$C = \frac{32}{F_{\text{MHz}}} \text{ meters}$$

$$D = \frac{8.8}{F_{\text{MHz}}} \text{ meters}$$

The tap point on the stub represented by these lengths will provide a good match to 300 ohm twin-lead, or to 75 ohm coaxial cable if a 4:1 BALUN transformer is provided.

The extended folded dipole shown in Figure 6.10 uses 300 ohm twin-lead for the folded dipole section (A) and single conductor wire for the vertical extensions (B). The length overall of section A is found from

$$A = \frac{123}{F_{\text{MHz}}} \text{ meters}$$

While the vertical sections are

$$B = \frac{9.2}{F_{\text{MHz}}} \text{ meters}$$

The antenna of Figure 6.10 can be fed with 300 ohm twin-lead, but this requires the antenna tuning unit to have a balanced output. Alternatively, you can feed it with 75 ohm coaxial cable if a 4:1 BALUN transformer is provided at the interface between the coaxial cable and the antenna feedpoint.

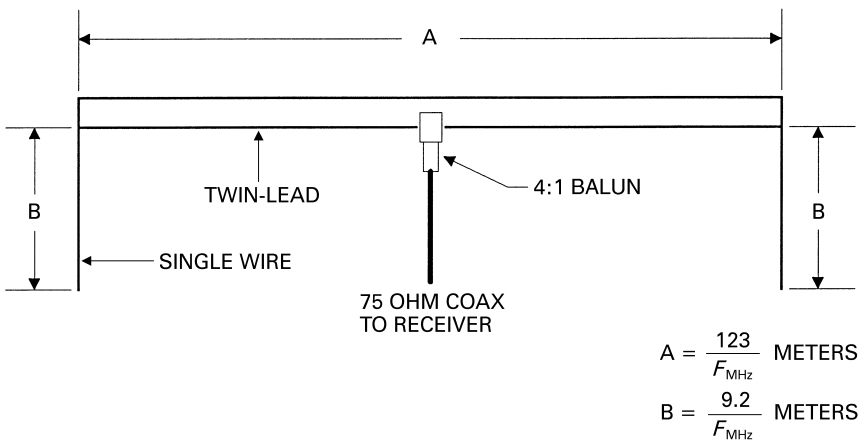


FIGURE 6.10

Limited Space Radial Layout

Radials are quarter-wavelength (usually) pieces of wire connected to the ground side of a transmission line at the antenna. These radials can be on the surface, under the surface, or elevated, depending on the particular antenna. The radials are used to form a **counterpoise ground**, i.e. an artificial ground plane. It is seen by the antenna as essentially the same as the ground.

By the way, the preferred 'ground level' configuration is buried, rather than on the surface, because it prevents injury to people passing by, either from RF burns or stumbling over the fool wire. But what do you do in the limited space situation?

A representative solution is shown in Figure 6.11. The radials shown in textbooks are straight, and that is the preferred configuration. But if you do not have the space, then you need to use some variant of the two paths shown in Figure 6.11. The radial can either go around the perimeter of the property, or zig-zag back and forth in either a triangle or rectangle pattern (the latter is shown here). I have even tacked radials to the baseboard of a student boarding house room (not recommended over a few watts of QRP power levels).

Imagination, a bit of engineering, some science, and a whole lot of luck make radio operations from limited space situations possible – or at least a lot easier.

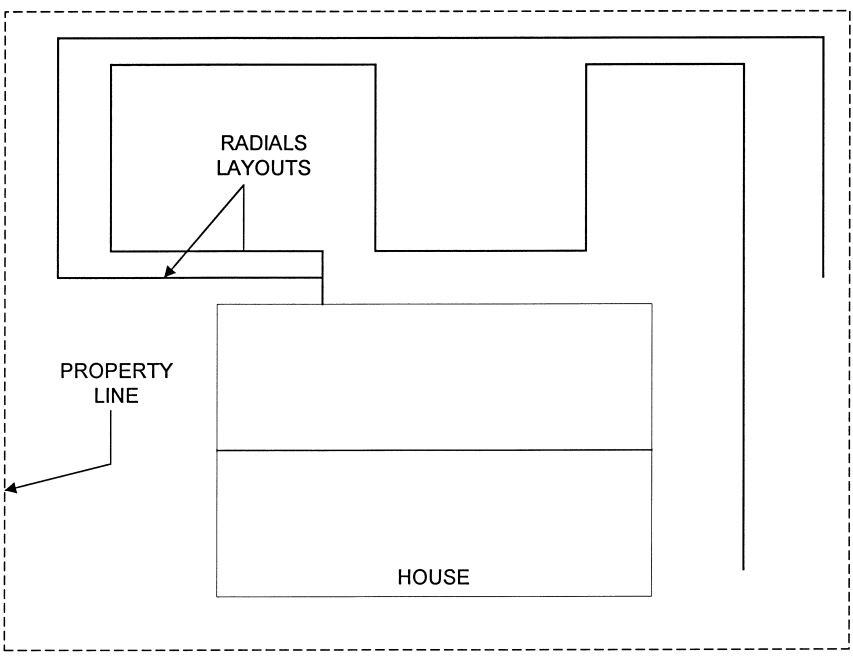


FIGURE 6.11

CHAPTER 7

Large loop antennas

There are two basic classes of loop antennas: **small loops** and **large loops**. The small loop antenna has an overall length that is less than about 0.18λ. Large loop antennas, on the other hand, have overall lengths larger than 0.5λ, and some of them are two or more wavelengths. The small loops are used for radio direction finding and for certain AM broadcast band reception problems. The topic of this chapter is the large loop antenna, of which several varieties for both transmitting and receiving are covered. We will also look at some in-between size loops that I have dubbed 'smaller large loops' to distinguish them from small radio direction finding loops.

QUAD LOOPS

The quad loop antenna (Figure 7.1) is perhaps the most effective and efficient of the large loop antennas, and it is certainly the most popular. The quad loop consists of a one wavelength loop of wire formed into a square shape. It provides about 2 dB gain over a dipole. The views in Figure 7.1 are from the horizontal perspective looking at the broad side of the loop. The azimuthal radiation pattern is a figure '8', like a dipole, with the directivity in and out of the page.

The quad loop can be fed in either of two ways. Figure 7.1A shows the feed attached to the bottom wire segment, and this produces horizontal polarization. The same polarization occurs if the feedpoint is in the top

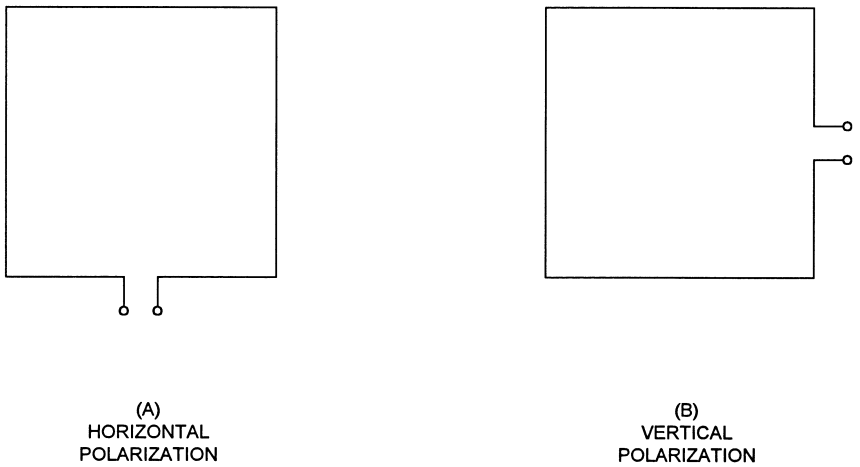


FIGURE 7.1

horizontal segment. If the feed is in either vertical segment (Figure 7.1B) then the polarization will be vertical.

The elevation pattern is shown in Figure 7.2. This pattern is found when the top horizontal segment of the loop is quarter wavelength from the ground. The directivity is in and out of the page. Note that there are two sets of lobes, one horizontally polarized (a minor lobe) and two vertically polarized. These lobes are derived by taking a slice out of the three-dimensional pattern that would be seen as a figure '8' pattern from above.

The overall length of the wire used to make the loop is found from

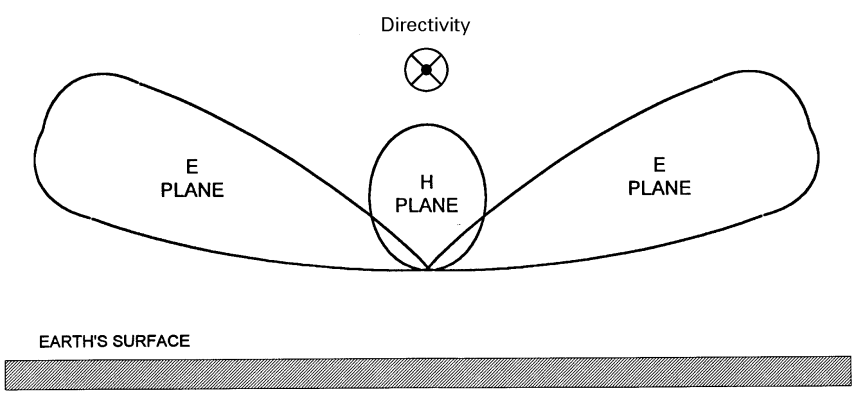


FIGURE 7.2

$$L_{\text{meters}} = \frac{306.4}{F_{\text{MHz}}} \text{ meters}$$

Examples of wire lengths for loops of various frequencies are given in Table 7.1. Each side is one-fourth of the total wire length.

There are several methods for constructing the quad loop. If you want a fixed loop, then it can be suspended from insulators and ropes from convenient support structures (tree, mast, roof of a building).

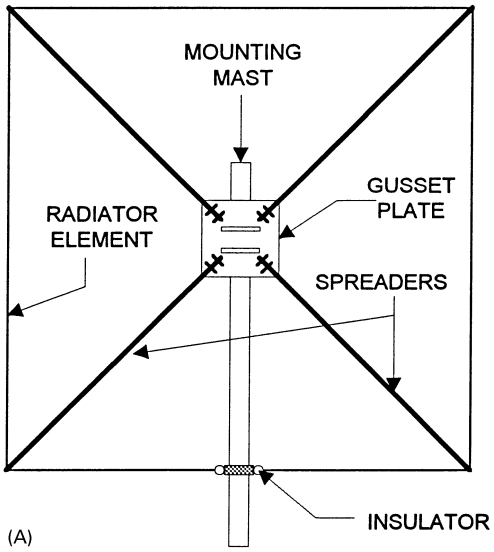
If you want to make the loop rotatable, then use a construction method similar to that of Figure 7.3A. This method uses a plywood **gusset plate**, approximately 30 cm square, to support a set of four **spreaders**. The gusset plate can be attached to a mounting pole or rotator pole with U-bolt clamps. Details of the gusset plate construction are shown in Figure 7.3B. The four spreaders are held to the gusset plate with two to four small U-bolts. The gusset plate is held to the mounting mast with two or three large U-bolts. In both cases, be sure to use a substantial size U-bolt in order to prevent breakage (use stainless steel wherever possible).

The spreaders can be made of wooden dowels at very high frequency (VHF). I have even seen larger cylindrical wooden dowels (2–3 cm diameter) used for quad loops in the upper end of the HF spectrum. For all other HF regions, however, you can buy fiberglass spreader specially manufactured for the purpose of building quad loops or quad beams. At one time, it was popular to build the quad loops from bamboo stalks. These were used as the core on which carpet was rolled, and carpet dealers would sell them for a modest price or even give them away. Today, however, carpet manufacturers use hard cardboard tubes for the roller, and these are unsuitable for building quad loops. Bamboo stalks of the right size (2.5–4 m) have all but disappeared from the marketplace.

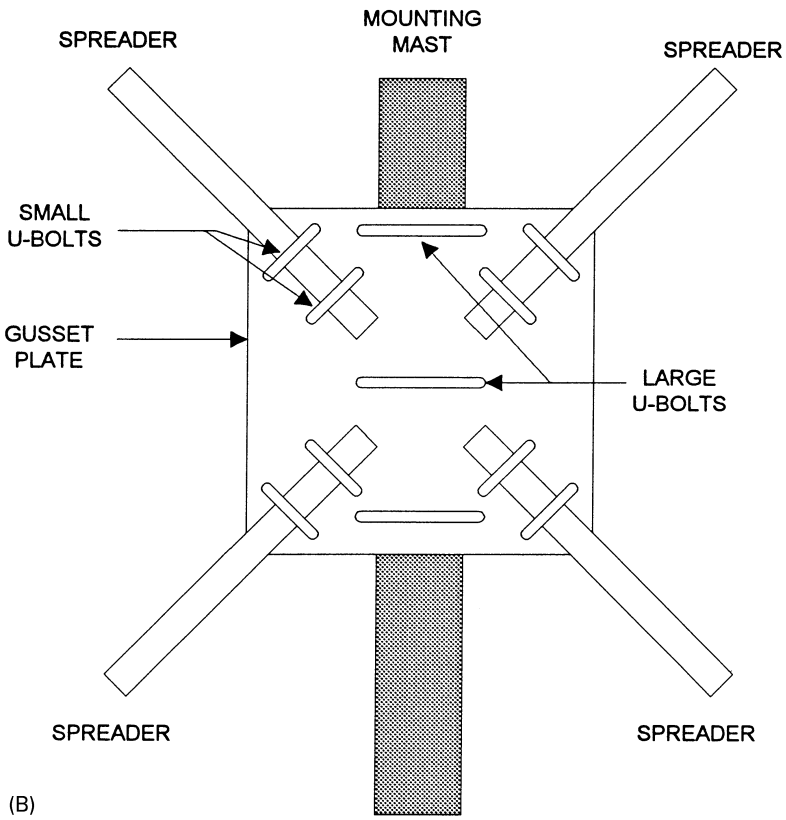
The quad loop can be fed with coaxial cable, although it is a good idea to use a 1:1 BALUN transformer at the feedpoint if only coaxial cable is used. The impedance match is not exact, and a voltage standing wave ratio

TABLE 7.1

Center frequency (MHz)	*L* (overall length) (m)	*L*/4 (each side) (m)
3.75	81.70	20.43
5.00	61.28	15.32
7.20	42.56	10.63
9.75	31.43	7.86
14.20	21.58	5.39
21.30	14.39	3.59
24.50	12.50	3.13
29.00	10.57	2.64


MOUNTING
MAST
GUSSET
PLATE
RADIATOR
ELEMENT
SPREADERS
INSULATOR

(A)


SPREADER
MOUNTING
MAST
SPREADER
SMALL
U-BOLTS
GUSSET
PLATE
LARGE
U-BOLTS
SPREADER
SPREADER

(B)

FIGURE 7.3

(VSWR) will be found. The feedpoint impedance is of the order of 100 ohms, so the VSWR when 75 ohm coaxial cable is used is only 100/75, or 1.33:1. Some people use a coaxial cable impedance matching stub between the quad loop feedpoint and the coaxial cable to the rig or receiver. Such a stub, called a **Q-section**, is made of 75 ohm coaxial cable, and is a quarter wavelength long. The coaxial cable to the rig or the receiver in that case is of 52 ohms.

To make an electrical quarter-wavelength matching stub you must shorten the physical length by the value of the velocity factor of the coaxial cable used for the matching section. For example, suppose we are building a quad loop for 14.25 MHz. A quarter wavelength in free space is $75/14.25 = 5.26$ m. But if polythene dielectric coaxial cable (velocity factor (VF) = 0.66) is used, the physical length required to make an electrical quarter wavelength is $5.26\,\text{m} \times 0.66 = 3.472$ m. If polyfoam dielectric coaxial cable (VF = 0.82) is used for the matching section, then the required physical length is 4.31 m.

QUAD LOOP BEAMS

The basic quad loop is bidirectional (figure '8' pattern). If you want to make the loop unidirectional, it can be built into a beam antenna by adding a second loop as either a **reflector** (Figure 7.4A) or **director** (Figure 7.4B). The reflector and director elements are not directly excited by the transmitter, and are therefore called **parasitic elements**. The reflector is slightly larger than the main loop (called the **driven element**), while the director is slightly smaller. The overall wire sizes are found from

Reflector:

$$L_{\text{Ref}} = \frac{315}{F_{\text{MHz}}} \text{ meters}$$

Driven element:

$$L_{\text{DE}} = \frac{306}{F_{\text{MHz}}} \text{ meters}$$

Director:

$$L_{\text{Dir}} = \frac{297}{F_{\text{MHz}}} \text{ meters}$$

Spacing between elements:

$$S = \frac{60}{F_{\text{MHz}}} \text{ meters}$$

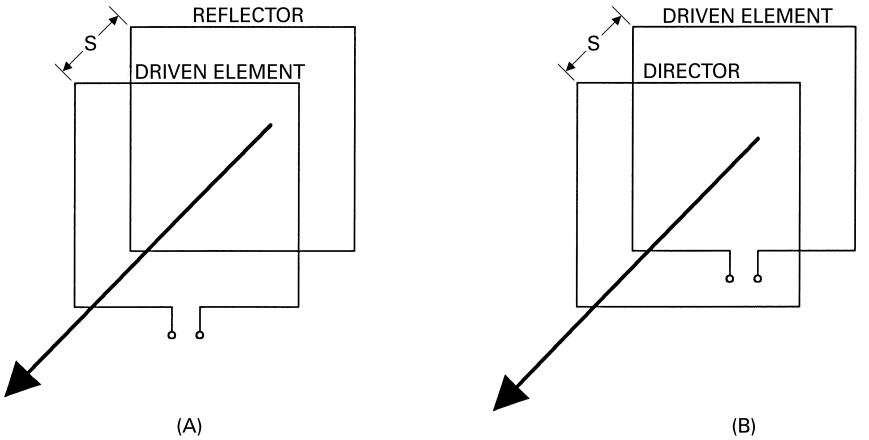


FIGURE 7.4

The directivity of the two-loop beam is always in the direction of the smallest element, as seen by the arrows in Figure 7.4, and the forward gain will be about 7 dBd (9.1 dBi), while the front-to-back ratio will be up to 25 dB (the actual value depends on the spacing of the elements).

The quad loop beam has a tradition that goes back to just before World War II. The missionary short-wave radio station HCJB in Quito, Ecuador was experiencing problems with its Yagi beam antennas. There was a constant corona discharge off the ends of the elements because the tips are high-voltage points. At lower altitudes than Quito's Andes location, these antennas do not exhibit the problem, but at altitude the arcing was severe enough to create a constant (and expensive) maintenance problem for the engineers. They invented the quad beam to solve the problem. The reason is that the feed method puts the high-voltage nodes in the middle of the vertical segments. It takes a much higher voltage to cause corona arcing from the middle of a cylindrical conductor than off the ends, so the problem was eliminated.

Ham operators and some commercial stations quickly picked up on the quad beam because it offered a relatively low-cost method for obtaining directivity and gain. In addition, the quad beam is believed to work better than the Yagi beam in installations that are close to the Earth's surface (e.g. less than a half wavelength).

The quad antennas shown here are two-element designs. Three and more elements can be accommodated by using both a reflector and a director, or any combination of multiple directors or reflectors. It is common practice (but by no means totally necessary) to use one reflector and as many directors as are needed to accomplish the desired number of

elements. Each element added to the array will increase the gain and narrow directivity. The feedpoint impedance is around 60 ohms, so it makes a reasonably good impedance match to both 52 and 75 ohm coaxial cable.

DELTA LOOPS

The delta loop (Figure 7.5) gets its name from the fact that its triangular shape resembles the upper-case Greek letter *delta* (Δ). These three-sided loops are made with a full-wavelength piece of wire, each side of the equilateral triangle being one-third wavelength long.

Three different feed schemes are shown in Figure 7.5, but all three of them attach the feedline at an apex of the triangle. The antenna in Figure 7.5A is fed at the top apex, that in Figure 7.5B is fed at bottom apex, while that in Figure 7.5C is fed at one of the side apexes (either right or left could be used).

The overall length (in meters) of the wire used to make the delta loop is found from $306.4/F_{\mathrm{MHz}}$ only if the antenna is mounted far enough from the ground and surrounding objects to simulate the elusive 'free space' ideal. But for practical delta loops a nearer approximation is found from $285/F_{\mathrm{MHz}}$. The actual size will be between the two values.

A delta loop built for 9.75 MHz is shown in Figure 7.6. The overall formula length of the wire is 29.23 m, with each side being 9.74 m. The transmission line to the rig or receiver is 52 ohm coaxial cable. There is a coaxial quarter-wavelength impedance matching section ('Q-section') between the antenna feedpoint and the line to the receiver or rig. The Q-section is made from 75 ohm coaxial cable.

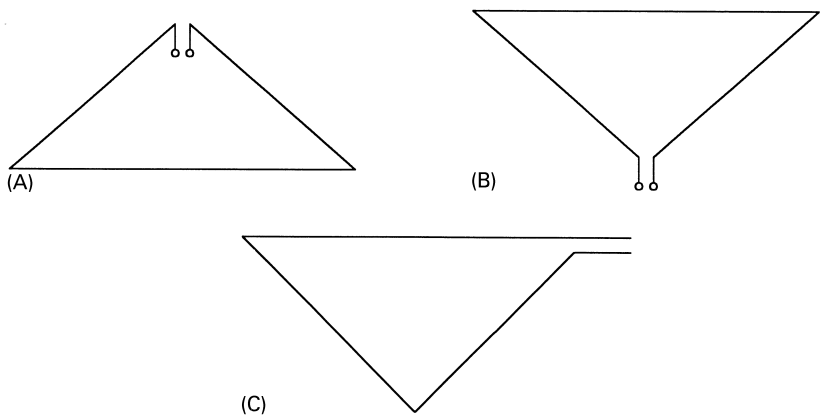


FIGURE 7.5

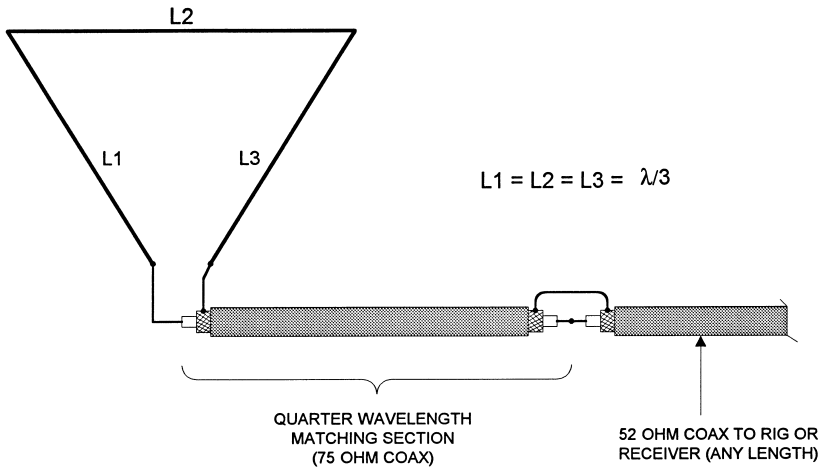


FIGURE 7.6

The physical length of the Q-section is reduced from the free space quarter wavelength ($75/F_{\mathrm{MHz}}$ m) by the VF, to ($75 \times \mathrm{VF})/F_{\mathrm{MHz}}$ m. At 9.75 MHz, if polythene coaxial cable is used (VF = 0.66) the length of the section is 5.1 m, but if polyfoam coaxial cable (VF = 0.82), then it is 6.3 m.

'ON-THE-CHEAP' ROOM LOOP

There are many situations where short-wave antennas cannot be installed without great pain and suffering, or where landlords, building managers, and other sundry unsympathetic types refuse to permit outdoor antennas. If you have a wall inside the building that will allow erection of a normal quad loop (as shown above), then there is no reason why you cannot attempt it (unless you are inside a metal frame or metal sided building!). I have seen cases where ham operators running relatively low-power levels (50–100 W) used full-wave quad loops mounted on the wall from floor to ceiling. In fact, one fellow sent me a photo of his installation where there were two adjacent quad loops on two walls that were at right angles to each other. By feeding them both along the vertical edge in the corner he was able to switch directions with a simple double-pole double-throw (DPDT) heavy duty switch. It seemed like a good solution for anyone who finds messy wire an acceptable wall decoration.

Another solution is to use a room loop antenna such as that in Figure 7.7. Although shown here as a horizontal delta loop, it can also be square or rectangular. The 'on-the-cheap' room loop is mounted on the ceiling of the room. The wire can be suspended from the type of hooks used to hang planters and flower pots. I have even seen cases where the sort of hooks

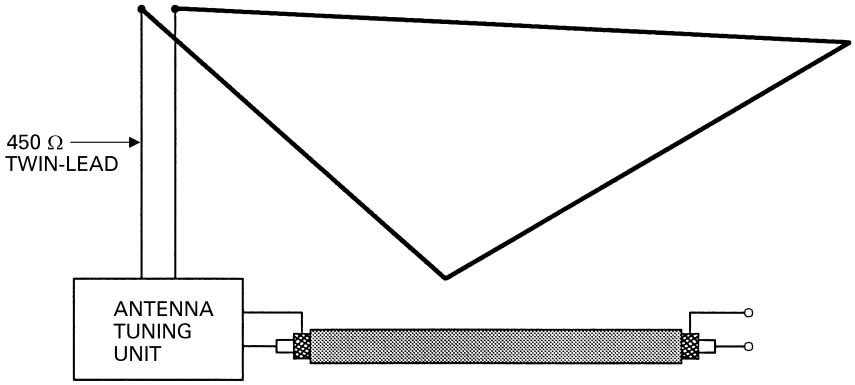


FIGURE 7.7

used to mount pictures on the walls were used on the ceiling to support the antenna wire.

The loop size is adjusted to fit the room, and is essentially a random length as far as any particular frequency is concerned. The feedline is 300 ohm, 450 ohm or parallel open feedline run vertically down the wall at the operating position to a balanced antenna tuning unit.

This antenna works well for receiving sites, and can be used for low-powered transmitters, but it should not be used for higher power transmitting. Also, it is probably prudent to use insulated wire for the antenna element.

INDOOR 'MIDDLE-EAST ROOM LOOP' ANTENNAS

There are many reasons why someone might want to build a room loop antenna. For some people, the problem is one of unsympathetic landlords, while in other situations there are problems in the layout of the property that limit antennas. In the USA, most new housing developments impose legally binding restrictive convenants that forbid the use of outdoor antennas on the deeds of houses sold. In some countries, short-wave radio antennas are limited because of political considerations. The antenna design in Figure 7.8 was sent to me via one of my magazine columns by a reader in a Middle-Eastern country that severely restricts both external antennas and the bands which short-wave receivers can pick up. He is limited to models that receive 4–7 MHz in addition to only a 200 kHz portion of the AM broadcast band and a 2 MHz segment of the FM broadcast band (what paranoia those politicians must have!). He also felt that it was necessary to not call attention to himself by erecting the short permitted outdoor antennas because, as a foreign worker, he is already somewhat suspect,

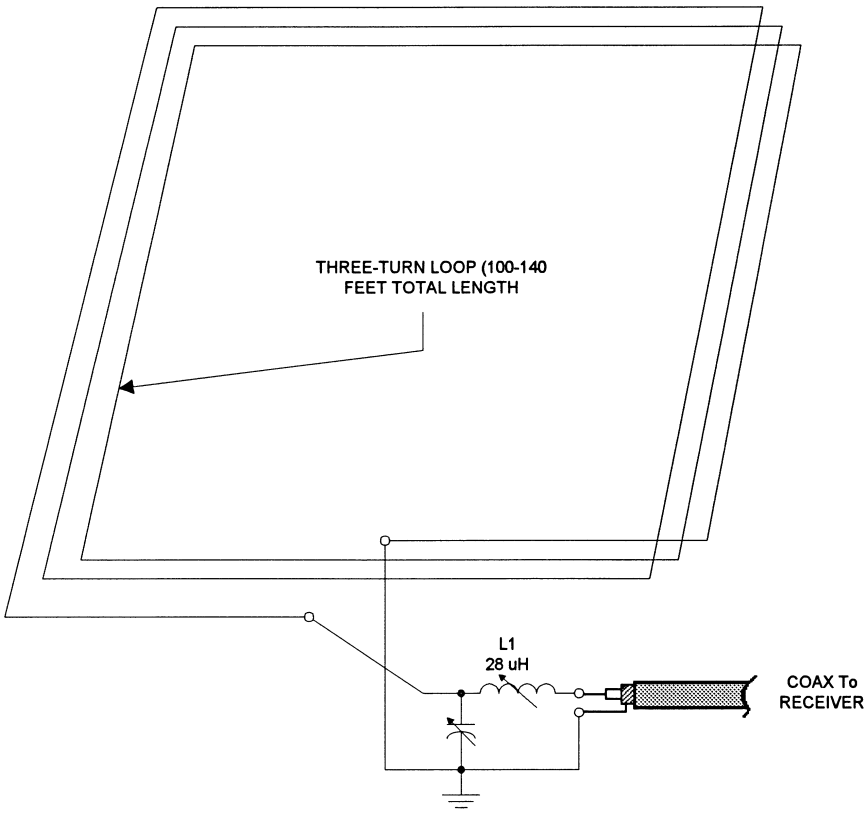


FIGURE 7.8

and believed that a substantial external antenna might label him as some sort of intelligence agent.

He claimed that this antenna worked quite well. The loop consists of 30–40 m of insulated wire (e.g. No. 22 hook-up wire) wound around the room to form three or four loops (the figure assumes a square room). My correspondent mounted the loop at the intersection of the ceiling and the walls. The windings were closely wound with little or no spacing between the windings.

The chap who sent me this design used an antenna tuner consisting of a shunt 365 pF broadcast variable and a 28 μH series inductor (i.e. a classic L-section tuner of the sort used by many radio amateurs and short-wave listeners). From some of my experiences during my student days (when I lived in student boarding house rooms), the antenna tuner might be purely optional for receiver operators. Try connecting your receiver directly to the loop using ordinary coaxial cable.

If you have some means for measuring the antenna impedance or VSWR, then make the measurement for the frequencies of interest before committing yourself to building or buying a tuner. Another option that will eliminate the antenna tuner, especially if you confine the use of the antenna to a small segment of the HF spectrum, is to use a BALUN or other form of impedance transformer at the feedpoint. This option becomes less viable as the frequency coverage required increases because the feedpoint impedance will vary all over the place with changes in frequency.

I cannot as yet recommend this antenna for transmitting above the purely QRP level (a few watts), so do not try it except at your own risk.

The directivity of this antenna is a bit uncertain because it will change with frequency. I found no problems using similar loops in my boarding house room, but common sense suggests that the azimuthal and elevation patterns will be quite different at the various bands throughout the HF short-wave bands.

A MULTIBAND SWITCHABLE DELTA LOOP

Wire antennas are among my favorites because they are low cost, reasonably well behaved, and can easily be erected and then torn down as the experimenting urge hits. Alternatively, if the urge to experiment is not something that hits you very often, or if you prudently lay down and let it pass when it does, then wire antennas are still (normally) easy to erect and are reasonably robust against the elements. They last a long time with reasonable maintenance. As might be expected of a book that deals with wire antennas, we take a look at quite a few different wire antennas. This time, we are going to look at two multiband delta loop antennas based on both the G4ABS and the N4PC designs. These antennas produce gains of the order of 4–8 dB, depending on the band of operation.

Multiband Loop 1

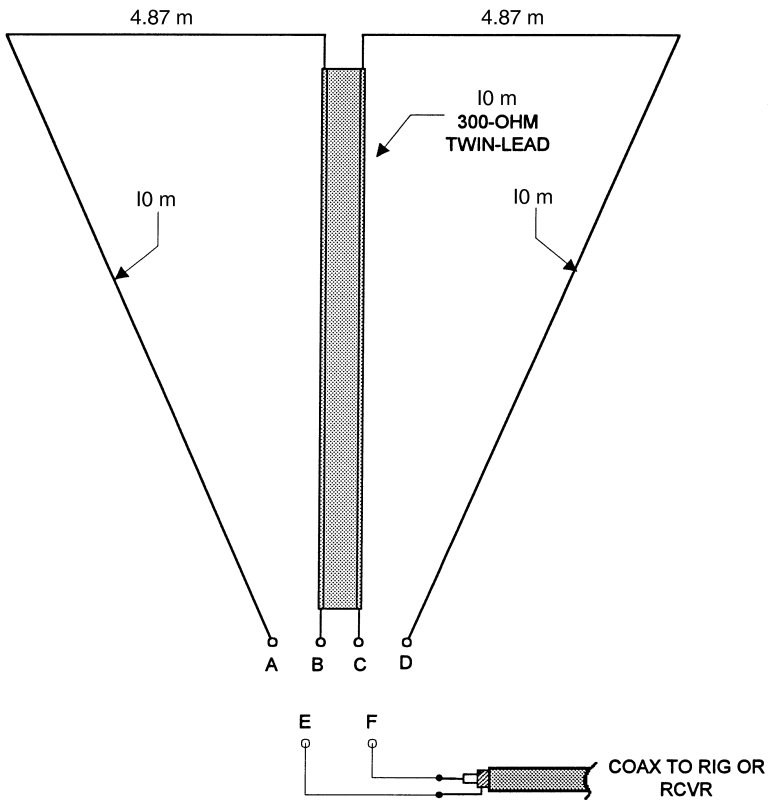
The first loop antenna works on 3.5, 7, 14 and 28 MHz, and is constructed of No. 14 antenna wire and a 10 m piece of either 300 or 450 ohm twin-lead transmission line. With slight modification, it will also work on a series of bands starting at 5 MHz and proceeding up through the HF short-wave band.

Figure 7.9A shows the multiband delta loop antenna design. There are four sections to this antenna. Two sloping side sections are made of No. 14 antenna wire, and are 10 m long. The top section is 9.75 m of No. 14 antenna wire, but is cut in the center to form two equal-length sections. Hanging from the center of the top section is a vertical section made of twin-lead transmission line (10 m).

To make this antenna work on the international short-wave bands rather than the ham bands, multiply these dimensions by 0.7. This is an example of dimensions scaling. It can be done by taking the ratio of the frequencies, and multiplying the resultant factor by the lengths. Scaling only works if all of the lengths are treated in this manner.

The two sloping side sections and the vertical section are brought to a connector board made of an insulating material such as ceramic, Plexiglas or even dry wood (if it can be kept dry and low power levels are used). We will take a look at the connection schemes for the different bands in a moment.

The antenna is supported in a manner similar to a dipole or any other wire antenna. The cut at the center of the top section, with its connections to the vertical section, are supported by an end insulator. The ends of the top half sections are supported by end insulators and ropes to nearby structures, just like a dipole. In some cases, a 10 m or higher wooden mast might be used to support the center of the top section. In that case, the vertical section



(A)

FIGURE 7.9

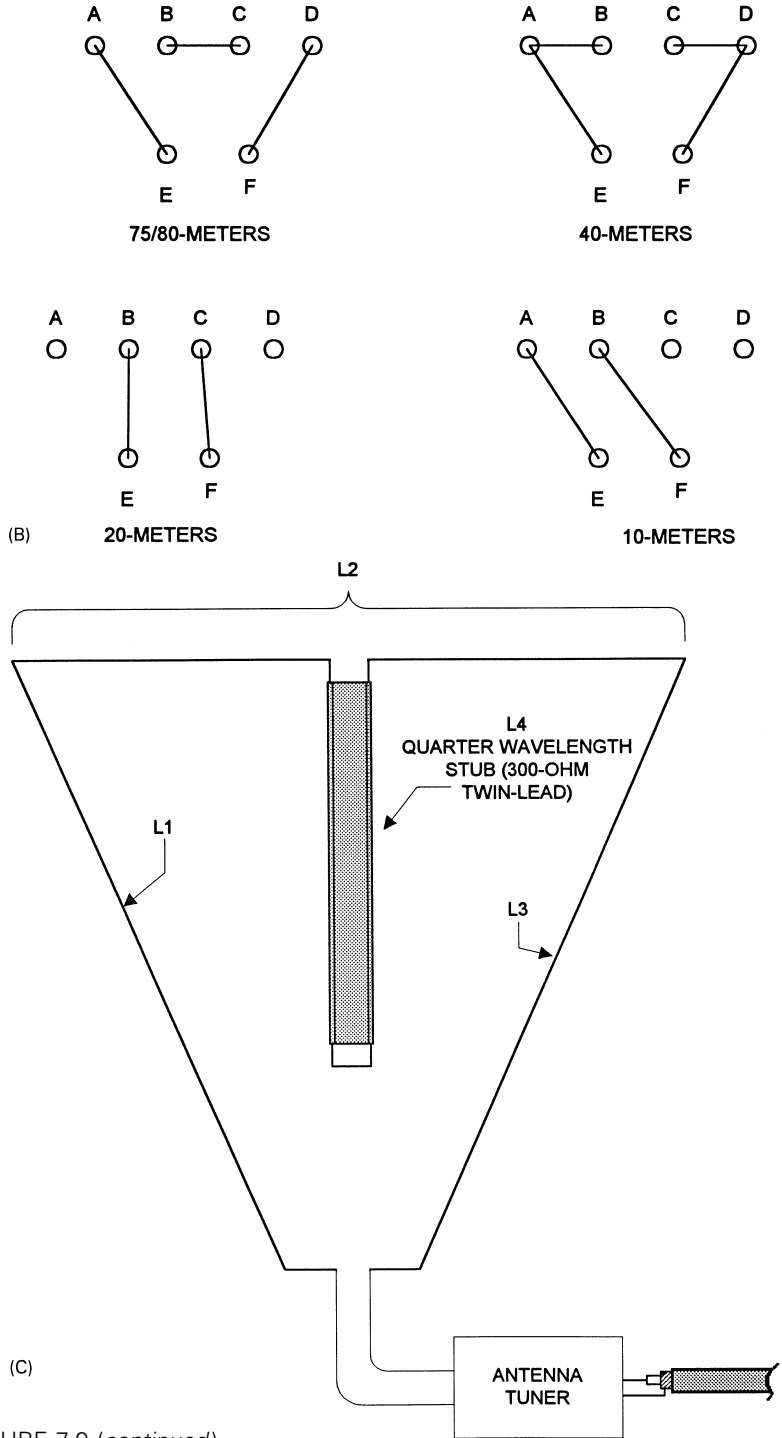


FIGURE 7.9 (*continued*)

can be tied to it, improving its mechanical support. In addition, the connections board (used for band changing) can be mounted close to its base.

Figure 7.9B shows the connections scheme to make the antenna work on the four different bands (which, you will undoubtedly notice are the harmonically related HF bands). All connections go to stand-off insulators, labeled A, B, C, D, E, and F. The cheapest way to use these connection points is to manually connect a shorting wire between the appropriate terminals. People with a bit more money to spend might want to get some coaxial relays, radio frequency relays, or vacuum relays in order to perform the switching. For receiver and very low-power transmitter applications, the connectors can be five-way binding posts that accept a banana plug. The shorting could be done with short (very short!) pieces of wire tipped at each end with the banana plugs. The connections scheme is as follows:

(1) To make the antenna work on the 75/80 m bands (3.5–4.0 MHz), short B–C, A–E, and D–F.
(2) To make the antenna work on 40 m (7.0–7.3 MHz), short B–A–E and C–D–F.
(3) To make the antenna work on 20 m (14.0–14.35 MHz), short B–E and C–F.
(4) To make the antenna work on 10 m (28.0–29.7 MHz) short A–E and B–F.

My impression of the multiband delta loop is that it works about as well as a dipole, and is generally bidirectional. However, the pattern does seem to change considerably from band to band (which is to be expected). I do not want to say much about how it changes because propagation effects are so different on the four bands that it would take a very long time of observing signals to get a good idea of what the pattern looks like. Any statement about patterns, other than it appears to change from band to band, would be unwise without extensive modeling or some actual measurements.

Multiband Loop 2

The other form of multiband loop is shown in Figure 7.9C. This design uses a full-wave loop with the balanced feedline connected at one of the apexes of the triangle, and a quarter-wavelength stub. Both the stub length and the overall length (L + L2 + L3) are calculated at the *highest* frequency of operation. It will operate over three bands that are harmonically related (e.g. 3.5/7.0/14.0 MHz, or 5/10/15 MHz, etc.). The overall length in meters is found from $(L1 + L2 + L3) = 615/F_{MHz}$, while the stub is $60/F_{MHz}$ (assuming that the usual form of twin-lead is used, which has a VF of 0.82). Both of these antennas produce patterns broadside from the loop

plane, although both the azimuthal and elevation radiation patterns vary from one band to another.

DOUBLE-DELTA LOOP

The double-delta loop antenna (Figure 7.10) consists of a back-to-back pair of large horizontal delta loops fed 180° out of phase with each other. The view shown is a plan view, i.e. as seen from above. The far corners of the two delta loops are connected together by a 180° phase-reversing harness made from a length of 450 ohm twin-lead or parallel ladder line. The 180° phase reversal is achieved by twisting the twin-lead over on itself once (and only once).

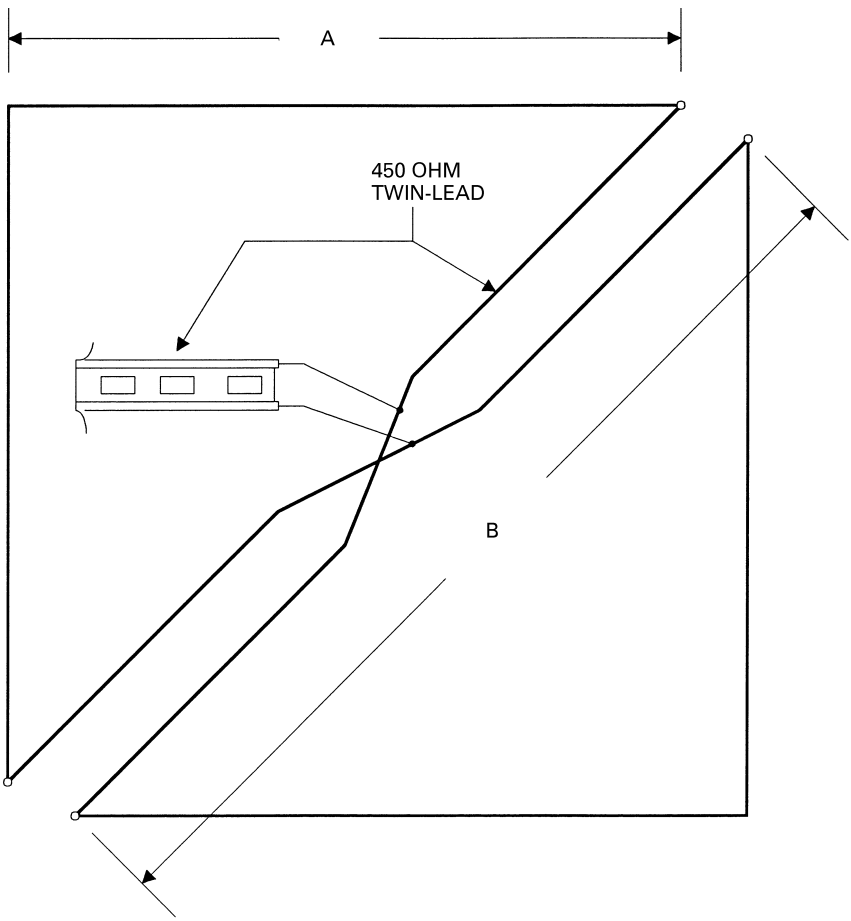


FIGURE 7.10

The feedline to the receiver or its antenna coupler is another piece of 450 ohm twin-lead connected to the mid-points of the two conductors of the 180° phasing harness. This transmission line will make a decent if not perfect match to receivers that have a balanced input. Use of a 4:1 BALUN transformer at the feedpoint will drop the impedance to about 112 ohms, which is not too bad a match for 75 ohm coaxial cable. On the other hand, if you want a near-perfect match, then opt instead for a 9:1 BALUN transformer and connect the receiver using 52 ohm coaxial cable. There are several forms of 9:1 BALUN transformer on the market, but are sometimes hard to find except by mail order. If such a transformer is not available, then you will have to build it yourself (which is not too great a chore).

The length of each of the four sides of the double-delta loop is about a quarter wavelength, and is found from

$$A = \frac{66}{F_{\text{MHz}}} \text{ meters}$$

Let us look at some examples: a 60 m band antenna cut for 5 MHz and a 31 m band antenna cut for 9.75 MHz (assume I did the arithmetic on my calculator, with the results tabulated below):

F (MHz)	Meters
5.00	13.20
9.75	6.77

The delta loop antenna is well regarded because it performs similarly to the quad loop, but is somewhat easier to install when the site only has a single high point. A friend of mine in Ireland uses a tall tower for his 20 m beam antenna. The tower becomes the upper support for a 75/80 m delta loop made of wire.

SMALLER LARGE LOOPS

The loops in this section are physically smaller than the one-wavelength loops of the previous sections. Few of them will work as well as the full-size loops, but they will work passably well, and can be installed at sites that will not accommodate the larger variety. I call them 'smaller large loops' to distinguish them from the small loops used in radio direction finding.

Half-wavelength loop

The half-wavelength loop is $\lambda/8$ on each side, and is one-half the size of the full-wavelength quad loop discussed earlier. There are two configurations for this loop. The closed-loop version (Figure 7.11A) transmits and receives in the direction opposite the feedpoint, while the open-loop variety (Figure 7.11B) transmits and receives in the oppposite direction.

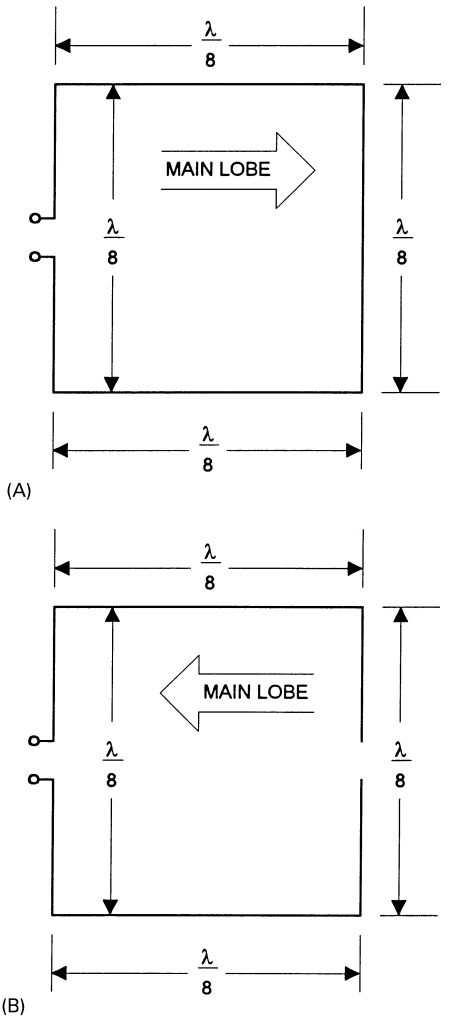


FIGURE 7.11

The two half-wavelength loops differ not just in the direction of radiation but also in the feedpoint impedance. The antenna in Figure 7.11A (closed-loop version) is fed at a voltage node, while the current node is in the middle of the opposite segment. As a result, the feedpoint impedance is several kilo-ohms. This means that a matching stub, impedance transformer or some other means of reducing the impedance is necessary. The open-loop variety (Figure 7.11B) forces the current to be a minimum at the open-circuit point, which means that the current maximum (and voltage minimum) occurs at the feedpoint. The open-loop version can be directly fed with coaxial cable.

These antennas are lossy, so have a poor front-to-back ratio (≈6 dB) and no gain over a dipole. Indeed, the gain is −1 dBd. Of course, if you want to make it seem like a better antenna, then quote the gain over isotropic, which is about +1.1 dBi.

Inductor-loaded smaller large loop

Figure 7.12A shows a less-than-full-size loop that is inductively loaded to lower the resonant point. The overall length of the wire used to form the loop is found from

$$A = \frac{180}{F_{\text{MHz}}} \text{ meters}$$

Each side is one-fourth this length. Note that the loop is a bit larger than a half wavelength.

The inductor should have a reactance of about 2500 ohms at the frequency in the middle of the band of operation. For example, if you build the antenna for reception in the 31 m band (e.g. 9.75 MHz), then the overall length of the antenna is 180/9.75 = 18.46 m, and each side is 4.62 m long.

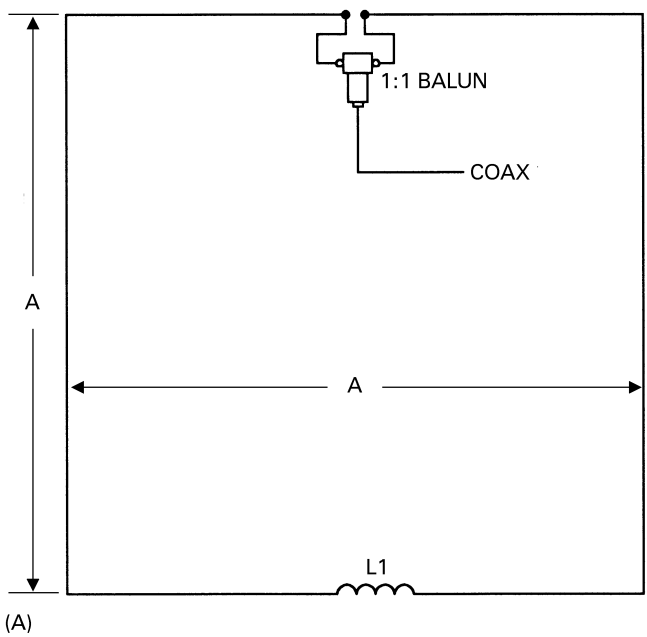
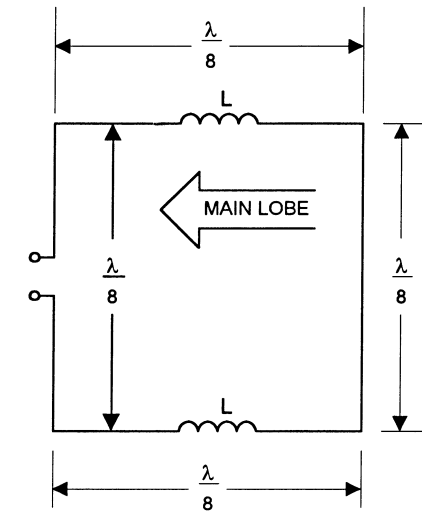
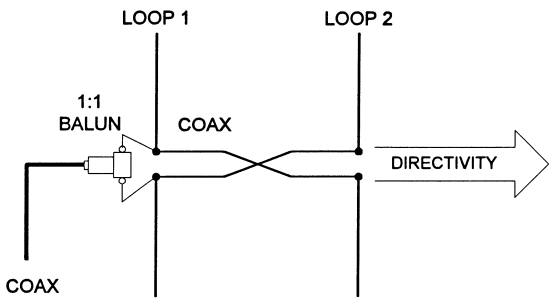


FIGURE 7.12



(B)



(TOP VIEW)

(C)

FIGURE 7.12 (*continued*)

The coil inductance required can be found from:

$$L_{\mu H} = \frac{400}{F_{MHz}} \text{ microhenrys}$$

For the 9.75 MHz example antenna, the value of the inductance required is (400/9.75 MHz) = 41 μH.

For receiver sites, the inductor can be a small toroid core coil. For example, if the 2.5 cm red core T-106-2 is used, then 55 turns of small diameter wire are required to make the inductor.

The antenna is not recommended for transmitting above the lowest power levels. Even at modest power levels (e.g. 100 W), a solenoid wound

air core coil should be of the order of 4–5 cm diameter with wire in the No. 14 size.

The antenna is fed at the mid-point of the side opposite the inductor. A 1:1 BALUN transformer is used to interface the feedpoint to 52 or 75 ohm coaxial cable.

A variation on the theme is the half-wavelength loop shown in Figure 7.12B. It is a tad smaller than the previous inductor load loop. In this case, the inductance is split into two inductors, and are placed at the mid-points of the segments adjacent to the segment containing the feedpoint. The coils in this variety should have a reactance of 360 ohms, making the inductance

$$L_{\mu H} = \frac{57}{F_{MHz}} \text{ microhenrys}$$

Both of these antennas can be used in a beam array by placing two identical loops facing each other (Figure 12.7C), and separated by a spacing of

$$S_{meters} = \frac{33}{F_{MHz}} \text{ meters}$$

The two antennas are fed in-phase, so there are two methods of attaching the transmission line. If you want to use parallel line or 450 ohm twin-lead, then connect the phasing harness straight across and connect the transmission line to the mid-point. But if you wish to use coaxial cable, then feed the antenna as shown in Figure 12.7C, using a 1:1 BALUN transformer. The 450 ohm twin-lead phasing harness between the elements should be twisted over on itself once, as shown.

The diamond loop

The diamond loop antenna is shown in Figure 7.13. This antenna is a shortened, flattened quad loop consisting of two equilateral triangles back to back. The length of each side, plus the height (dashed line) of the two principal apexes, is found from

$$L_{meters} = \frac{70}{F_{MHz}} \text{ meters}$$

In order to feed this antenna with coaxial cable, a 4:1 BALUN transformer is used between the coaxial cable and the feedpoint of the antenna.

Half-delta loop

The antenna in Figure 7.14 achieves a smaller size by being grounded. It is essentially one-half of a full-wave delta loop, with the other half being 'imaged' in the ground. The wire forms a right triangle, with the vertical

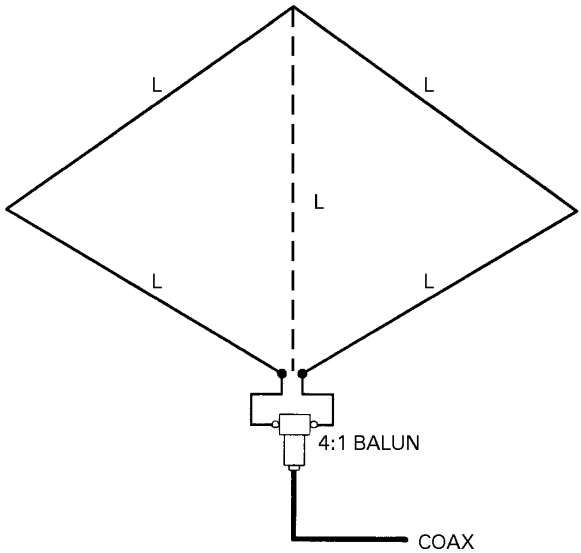


FIGURE 7.13

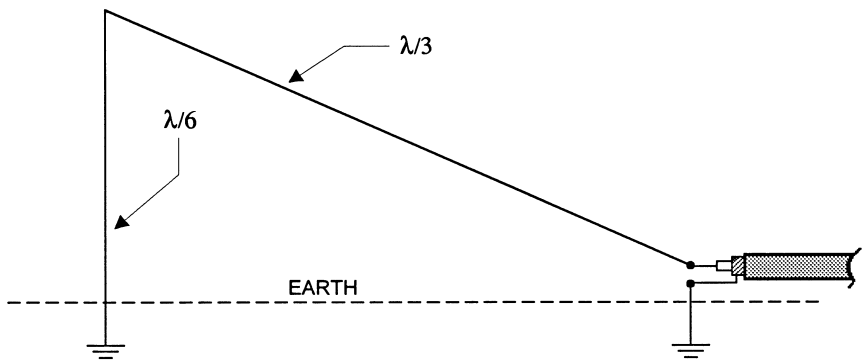


FIGURE 7.14

section being $\lambda/6$ and the sloping section being $\lambda/3$. These lengths are found from

$$\frac{\lambda}{3} = \frac{102}{F_{\text{MHz}}} \text{ meters}$$

and

$$\frac{\lambda}{6} = \frac{61}{F_{\text{MHz}}} \text{ meters}$$

The antenna is inherently unbalanced with respect to ground, so can be fed directly with 52 ohm coaxial cable.

Two-band compact loop

Most of the loops discussed thus far are basically monobanders, unless multiple loops are built on the same frame and fed in parallel. The loop in Figure 7.15, however, operates on two bands that are harmonically related to each other. For example, if FL is the lower band, and FH is the higher band, then FH = 2 × FL.

The overall length of the loop is a half wavelength, but it is arranged not into a square but rather into a rectangle in which the horizontal sides are twice as long as the vertical sides, i.e. the horizontal elements are a quarter wavelength and the vertical sections are one-eighth wavelength. The section lengths are

Horizontal:

$$L_{\text{hor}} = \frac{75}{F_{\text{MHz}}} \text{ meters}$$

Vertical:

$$L_{\text{ver}} = \frac{37.5}{F_{\text{MHz}}} \text{ meters}$$

In both equations, the frequency is the center frequency of the lower band of operation.

The vertical stub is made of 600 ohm parallel open-wire transmission line, although 450 ohm twin-lead could also be used. The length of the stub is found from the same equation (above) as the vertical segment if open-wire line is used. If twin-lead is used, then multiply that distance by the velocity factor of the transmission line.

The coaxial line is a Q-section made of 75 ohm transmission line. It is cut to a quarter wavelength of the upper band, i.e. twice F_{MHz} used in the calculations above.

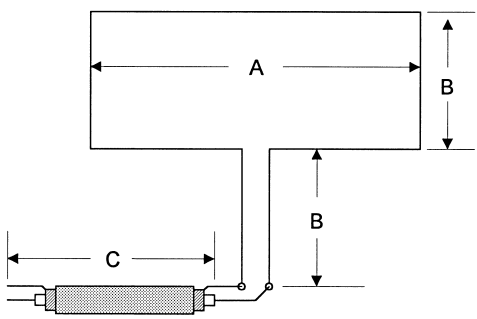


FIGURE 7.15

THE BI-SQUARE REALLY LARGE LOOP

The bi-square loop in Figure 7.16 is twice as large as the quad loop discussed earlier in this chapter. The overall length of the wire is two wavelengths, so each side is a half wavelength long. The overall length is calculated from

$$L_{\text{meters}} = \frac{585}{F_{\text{MHz}}} \text{ meters}$$

While each side is:

$$L_{\text{meters}} = \frac{146.25}{F_{\text{MHz}}} \text{ meters}$$

The bi-square antenna can be used on its design frequency, and also at one-half of its design frequency (although the patterns change). At the design frequency, the azimuthal pattern is a clover leaf perpendicular to the plane of the loop, and is horizontally polarized. At one-half the design frequency the radiation is vertically polarized, and the directivity is end fire.

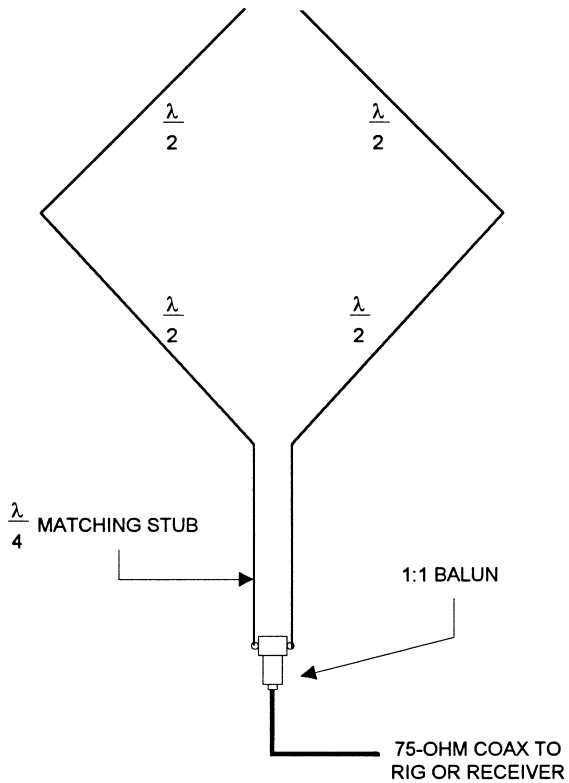


FIGURE 7.16

VHF/UHF QUAD BEAM ANTENNA

Figure 7.17 shows a VHF/UHF quad beam with four elements. Although it's technically possible to build such a beam for a lower frequency than VHF, it is not too practical because of mechanical considerations. But at VHF and UHF frequencies the dimensions of the elements become less large, and therefore more manageable. These dimensions are:

Reflector:

$$L_{\mathrm{REF}} = \frac{31\,500}{F_{\mathrm{MHz}}}\ \mathrm{cm}$$

Driven element:

$$L_{\mathrm{DE}} = \frac{30\,600}{F_{\mathrm{MHz}}}\ \mathrm{cm}$$

Directors:

$$L_{\mathrm{DIR}} = \frac{29\,700}{F_{\mathrm{MHz}}}\ \mathrm{cm}$$

Spacing between elements:

$$S = \frac{6000}{F_{\mathrm{MHz}}}\ \mathrm{cm}$$

Where all elements lengths are in centimeters (cm).

The antenna can be made of thin or thick wire. In the thin wire (#10 through #18) version, a support such as Figure 7.3 is needed. In the thick wire version, clothes line or other thick wire (#00 to #10) is needed. In that case, it might be possible to erect the antenna without any support other than the boom that holds all four elements.

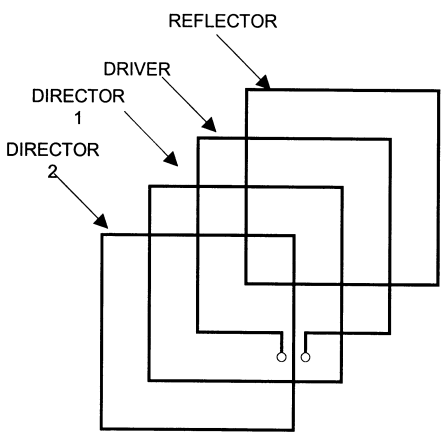


FIGURE 7.17

CHAPTER 8

Wire array antennas

The antennas found in this chapter are any of several array designs that use quarter-wavelength, half-wavelength, or other radiator elements arranged in a fixed array. There are several ways to categorize array antennas. A **collinear array** consists of two or more elements, ranging in length from 0.25λ to 0.65λ (some texts say 1λ), fed such that the currents in the elements are *in-phase* with each other.

Arrays are also classified according to the directivity: **broadside arrays** radiate and receive along a line perpendicular to the plane of the antenna, while **end-fire array**s radiat and receive off the ends of the elements.

The down side of these types of antennas is that many of them require parallel open-line transmission line (although some can use 300 or 450 ohm twin-lead), and impedance matching stubs. But we can handle that problem. Chapter 3 showed what parallel line and the two forms of twin-lead transmission line look like.

Some of the antennas can be fed directly with a 4:1 or 1:1 BALUN transformer and 75 ohm coaxial cable, but in many cases an **impedance matching stub** is needed (see Chapter 11). The BALUN transformers can either be homemade, or store-bought. Although antenna and scanner supplies distributors sell very high-frequency (VHF) BALUN transformers, you can also use television-style BALUN transformers. These may not be called BALUNs on the package, but are recognized by the fact that they have a coaxial 'F' connector at one end, and a piece of 300 ohm twin-lead at the other.

There are several antennas that fit either tightly or loosely into the general category of 'array antennas.' The selection of which antennas to put in this chapter, and which to put in Chapter 5 (dipoles, doublets), was not entirely clear in each case. This ambiguity was especially annoying when looking at antennas such as the double extended Zepp, which I put in Chapter 5. The problem is that antennas such as the double extended Zepp are quite properly called 'array' antennas even though they look suspiciously like doublets (which they are, of course). Other antennas were less difficult to place unambiguously in one chapter or the other.

ARRAY ANTENNAS

The idea in an array antenna is to provide directivity and gain by using two or more antenna elements in such a way that their fields combine and interact to focus the signal in one direction, or a limited number of directions (e.g. bidirectional, like a dipole). Figure 8.1A shows an array not unlike some of those in this chapter. It consists of two half-wavelength dipoles stacked one on top of another a distance (*S*) apart. The view here is broadside from the horizontal direction, so the dipole fed at A1–A2 is positioned above that which is fed at B1–B2.

One requirement of this form of antenna is that the two half-wavelength dipoles be driven in-phase with each other (some other antenna arrays want out-of-phase excitation). That means that the phase of the signal applied to A1–A2 is exactly the same as the phase of the signal applied to B1–B2. This is accomplished with either of the two methods shown in Figure 8.1B and

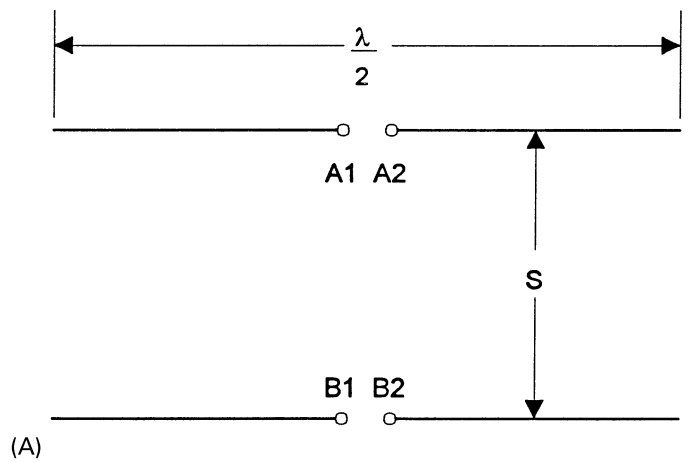
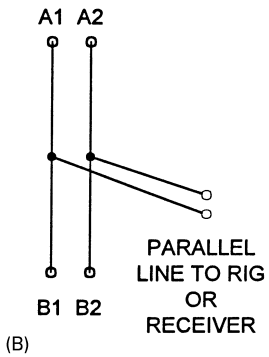
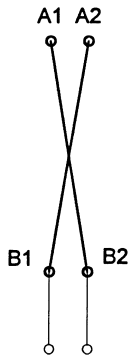


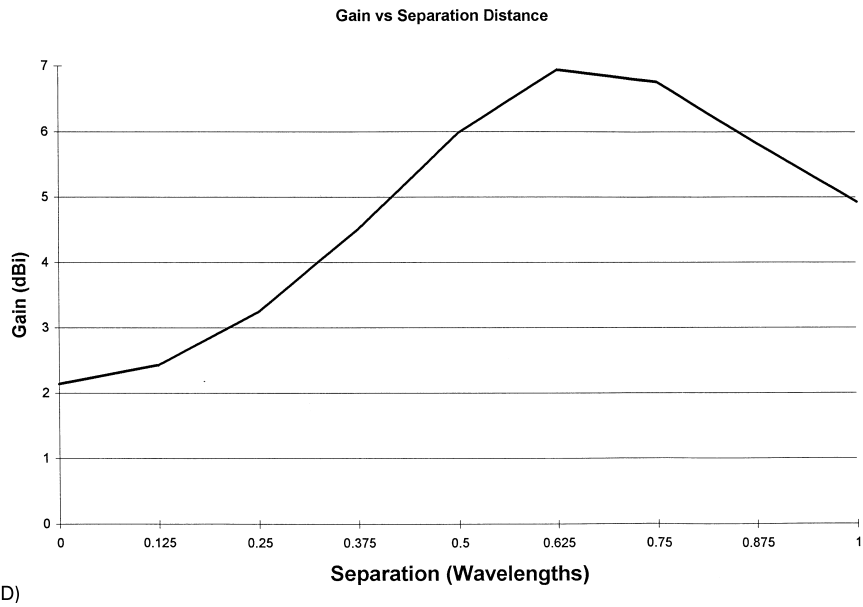
FIGURE 8.1



(B)



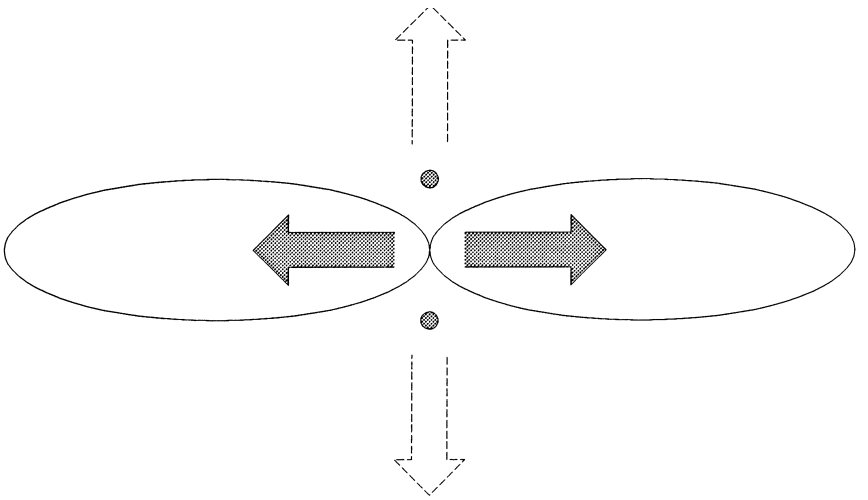
(C)



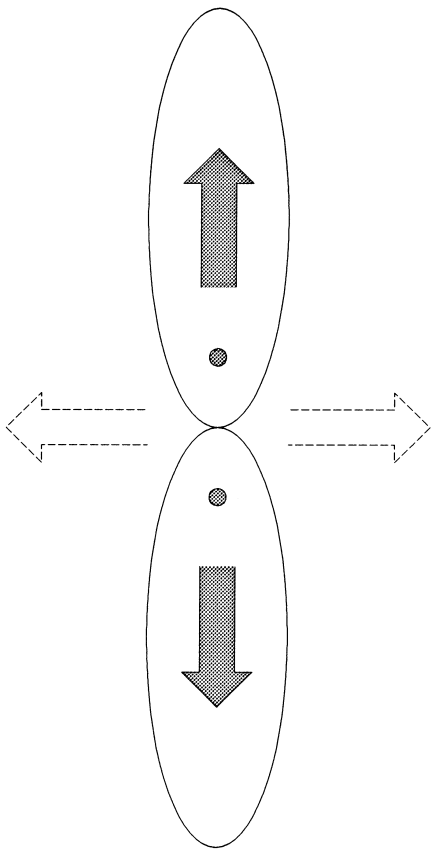
(D)

FIGURE 8.1 (*continued*)

8.1C. In Figure 8.1B, a piece of parallel transmission line is connected between the feedpoints such that A1 is connected to B1, and A2 is connected to B2. Another piece of parallel feedline from the receiver or transmitter is connected at the exact mid-point on the harness between the two dipoles.



(E)



(F)

FIGURE 8.1 (*continued*)

When this is done correctly, the voltages appearing across A1–A2 are exactly the same amplitude and phase as the voltages across B1–B2.

In the method used in Figure 8.1C, the parallel line from the rig or receiver is connected to the feedpoint of one dipole (in this case B1–B2). In order to preserve the 180° phase relationship between the elements, the feedline phasing harness between A1–A2 and B1–B2 must be twisted once such that A1 is connected to B2, and A2 is connected to B1.

The gain realized by stacking the two half-wavelength dipoles in this manner is a function of the spacing (S) between the elements. The approximate gain above isotropic that can be realized is shown in Figure 8.1D (note: to convert to gain above a dipole, subtract 2.14 dB from each value). Note that the gain peaks at about 6.94 dBi (or 4.8 dBd) in the vicinity of $5\lambda/8$ (0.625λ). This is the maximum gain realizable from this configuration. Although it looks good, there are sometimes other factors that deteriorate gain performance, so some experimentation with element spacing is usually in order for those who desire peak performance.

The antenna pattern produced by phased array antennas depends on the nature of the elements, the spacing between them and the ratio of the currents flowing in the elements. In the simple case of two elements spaced a half wavelength apart, we will see the general patterns of Figures 8.1E and 8.1F. These patterns are azimuthal plots as seen from above; the antennas are omnidirectional vertical radiators. The pattern in Figure 8.1E is for the case where the two antenna elements are fed in-phase with each other. The maxima are perpendicular to a line between the two antennas, while the minima are off the ends. Feeding the antennas out-of-phase with each other (Figure 8.1F) flops the pattern over so that the maxima are along the line between the elements, and the minima are perpendicular to this line.

There are a large number of patterns for various spacings and phase angles. The originals were developed in 1929 by a chap named Brown, and published in the USA in *The Proceedings of the Institute of Radio Engineers* (now the Institute of Electrical and Electronics Engineers). Brown's patterns are reproduced in *The ARRL Antenna Handbook* and any of several antenna engineering handbooks. Now let us look at some antennas.

TWO-ELEMENT COLLINEAR ARRAY

One of the simplest forms of gain array antenna is the two-element collinear array (shown in Figure 8.2 as a side view). It consists of two half-wavelength dipoles spaced a half wavelength apart. This antenna gives a gain of 3 dB, and is a broadside array (when viewed from above the pattern is perpendicular to the line between the antenna elements).

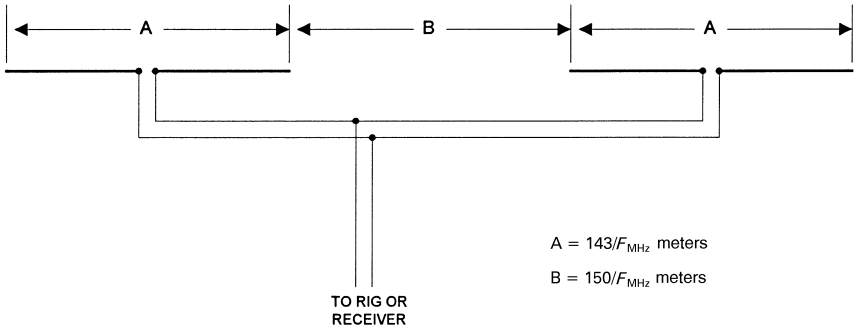


FIGURE 8.2

Each antenna element is a half wavelength, but the actual lengths and the equations used to calculate them are a little different for the HF and VHF/UHF bands.

HF band lengths

On the HF bands (3–30 MHz) the antenna is most likely mounted close to the Earth's surface, so there will be some end-effect reduction of the antenna length due to capacitive coupling. There is also a small reduction (2–3%) due to the velocity factor of the signal in the wire (this is set by the length/diameter ratio of the wire). As a result, the length of the elements is typically reduced 4–5% compared with the free space length. The gap between the dipoles, however, is calculated at the free space value. Thus,

$$A = \frac{143}{F_{\text{MHz}}} \text{ meters}$$

and

$$B = \frac{150}{F_{\text{MHz}}} \text{ meters}$$

Typical lengths for popular bands are given in Table 8.1.

TABLE 8.1

Band (MHz)	*A* (m)	*B* (m)
7.00	20.40	21.43
9.75	14.63	15.40
14.20	10.06	10.55
16.00	8.93	9.39
21.25	6.70	7.07
24.40	5.85	6.16

VHF/UHF band lengths

The physical length of the dipoles for VHF can be calculated from the free space version of the equation because these antennas are typically installed a sufficient number of wavelengths above the Earth's surface that the free space constant is reasonable. Thus, both A and B are the same physical length. These lengths are found as $A_{cm} = B_{cm} = 2324/F_{MHz}$, where F_{MHz} is the frequency in megahertz. Thus, at 150 MHz, the half wavelength is 15.49 cm; at 450 MHz it is 5.16 cm; and at 800 MHz it is 2.91 cm.

The antenna in Figure 8.2 is a little inconvenient in one respect: it requires two transmission lines connected in parallel. This is easily accomplished by using 75 ohm coaxial cable from each antenna element to a 'tee' connector, and then a piece of 52 ohm cable from the 'tee' connector to the receiver. At HF frequencies this antenna can take up a lot of 'footprint' space.

A different solution is shown in Figure 8.3. The antenna elements are separated by only a few centimeters, rather than a half wavelength. The gain drops to 1.6 dB for two elements (although for three elements it is 3 dB, and for four elements it is 4.2 dB). The antenna is fed by 450–600 ohm parallel line or twin-lead. The feedpoint impedance will be as low as 1 kilo-ohm if tubing is used, and up to 4–6 kilo-ohms if wire is used. This is one of those antennas where a matching stub is needed.

The four-element version of the antenna is shown in Figure 8.4 (it is called the **collinear array** or **Franklin array**). This antenna is similar to the two-element version of Figure 8.3, but with an additional half-wavelength element tacked onto the ends. The additional elements are separated by quarter-wavelength **phase reversal stubs**. The stubs are needed to keep the phase of the currents flowing in the outer elements the same as the currents flowing in the center elements. These stubs are made from the same type of

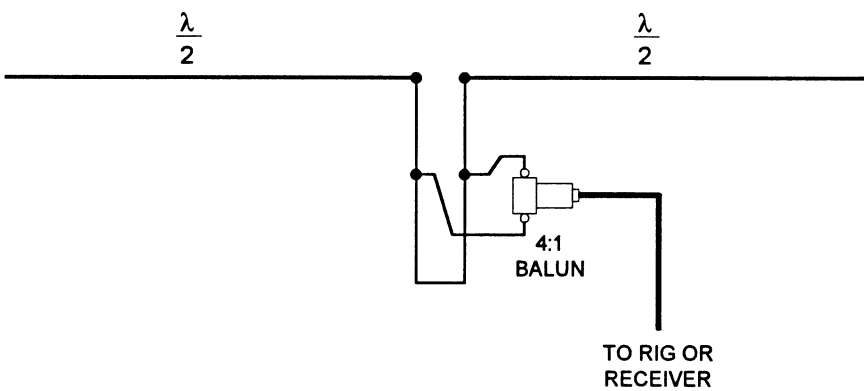


FIGURE 8.3

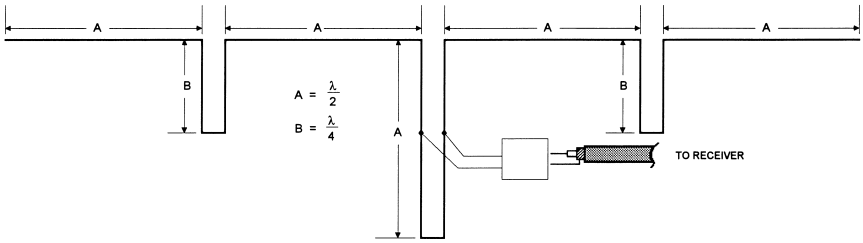


FIGURE 8.4

transmission line as the half-wavelength matching stubs, but are half as long.

Another collinear array is shown in Figure 8.5. This antenna offers 3 dB of gain, and can be fed directly from 75 ohm coaxial cable through a 4:1 BALUN transformer. The antenna consists of three half-wavelength elements, but the center element is fed at the center point (a quarter wavelength from each end). The quarter-wavelength phase reversal stubs are used between the center element and the two outer elements.

FOUR-ELEMENT BROADSIDE ARRAY

A four-element vertical broadside array is shown in Figure 8.6. Although often implemented using aluminum tubing, the array can easily be erected using wire elements as well. This antenna has four half-wavelength elements spaced a half wavelength from each other. Like the antenna above, the radiation direction is perpendicular to the line between the antennas, or in and out of the page.

The antenna elements are fed by a length of 600 ohm parallel line that connects all four together. But notice that the outside elements are fed by reversing the phase. This means that the transmission line must be twisted

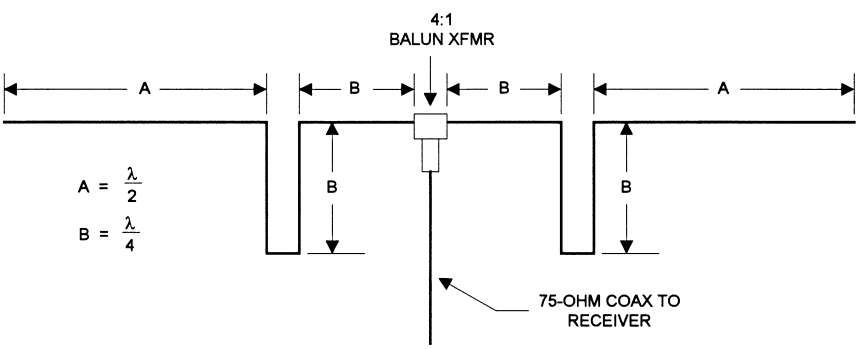


FIGURE 8.5

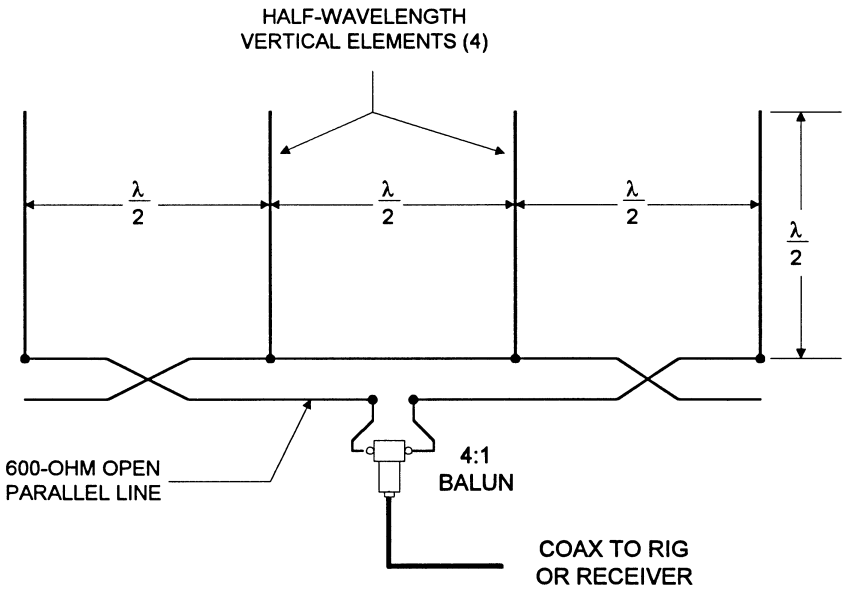


FIGURE 8.6

before being connected to the end elements; the unused conductor is left floating. The feedpoint impedance of this antenna is 200–300 ohms, so it can be fed with either 52 or 75 ohm coaxial cable if a 4:1 BALUN transformer is provided. Gains of the order of 6–8 dB are realized with this antenna.

EIGHT-ELEMENT BROADSIDE ARRAY

The eight-element broadside array shown in Figure 8.7 is built using similar methods to the previous antenna, but with two bays of elements rather than just one. The overall height of the array is one wavelength (although half-wavelength versions can also be built at some cost to gain). As in the previous case, the elements are spaced a half wavelength apart and are fed at their bases through 600 ohm parallel line that is twisted at the ends. Rather than breaking one conductor to connect the feed, this antenna connects the BALUN transformer across the antenna phasing harness wires. Note also that a 1:1 BALUN is used here, rather than a 4:1 device as above. This antenna is capable of as much as 12 dB gain.

THE LAZY-H ARRAY

The lazy-H array is a special case of the doublet array in which the doublets making up the array are one wavelength long (Figure 8.8A), fed in-phase

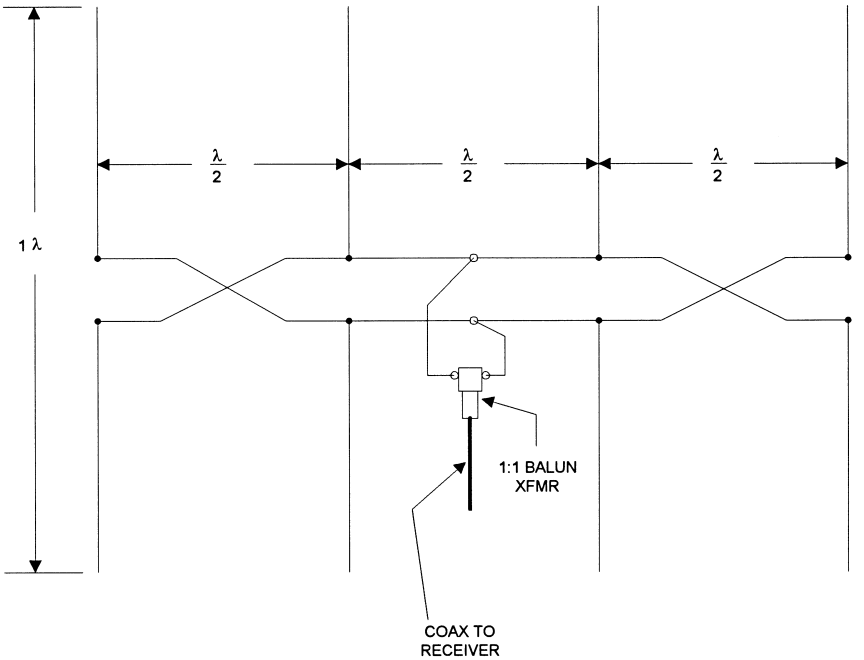


FIGURE 8.7

with each other. The directivity of this antenna is broadside, and so radiates in two directions into and out of the page. This view is as seen from the front or back, so note that the two one-wavelength doublets are stacked one on top of the other.

Gains for the lazy-H antenna vary with the spacing between the doublets (S), and will be as low as 4 dB at $3\lambda/8$ to almost 7 dB at $3\lambda/4$ spacing. The antenna should be installed at least a quarter wavelength above the Earth's surface.

The two doublets in the lazy-H must be fed in-phase with each other. When the feed arrangement is as shown in Figure 8.8A, i.e. with the feedpoint in the center, the transmission line phasing harness is connected straight across (no twisting). If only single-band operation is anticipated, then the feedpoint impedance will be a reasonable match to 75 ohm coaxial cable; a 1:1 BALUN transformer is used to connect the coaxial cable to the phasing harness. For multiband operation, however, it is necessary to use tuned feeders to an antenna tuning unit.

An alternative feed scheme is shown in Figure 8.8B. In this arrangement, the feedpoint is at the feedpoint of one of the doublets, rather than in the center of the harness. In order to preserve the in-phase relationship, the

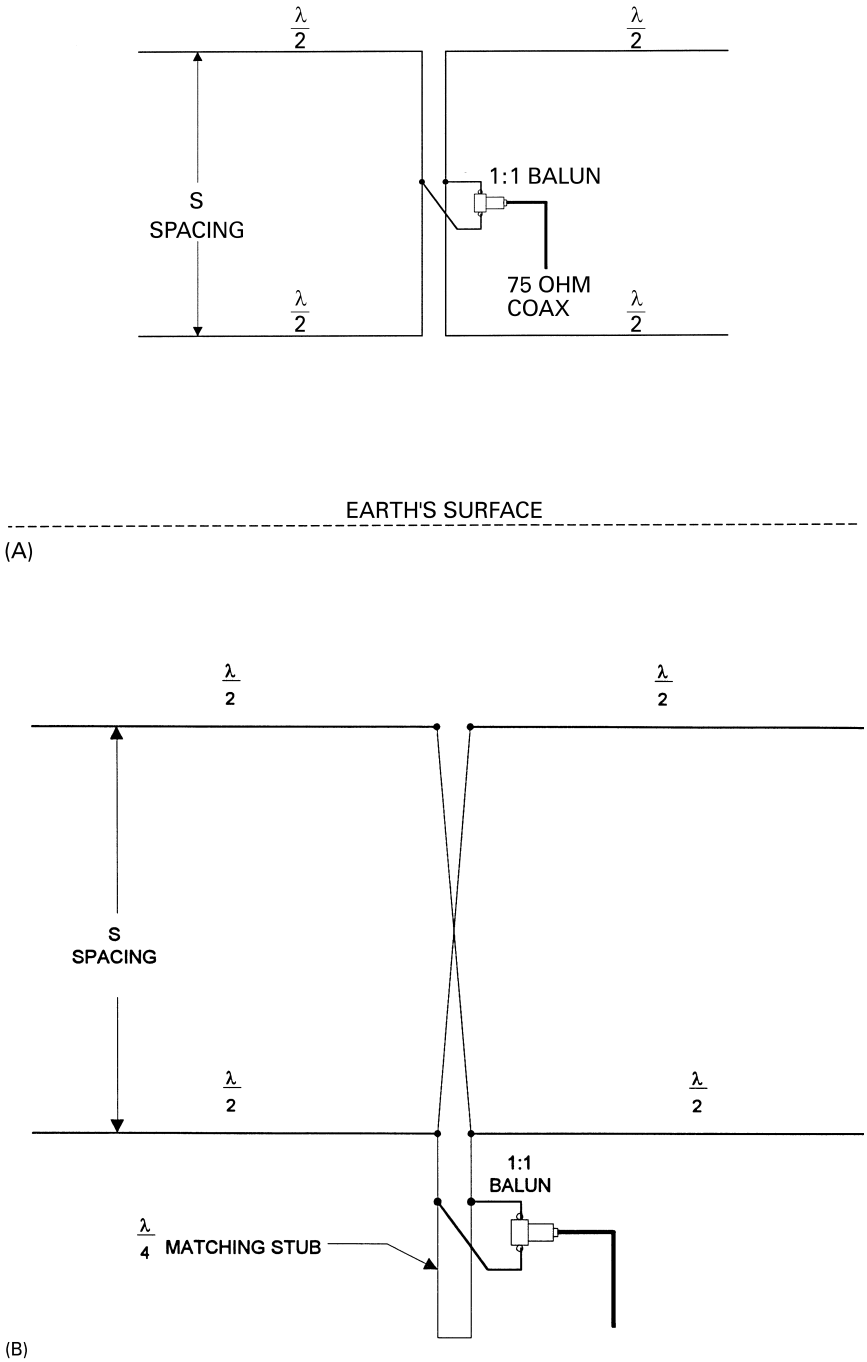

λ/2
λ/2
S
SPACING
1:1 BALUN
75 OHM
COAX
λ/2
λ/2
EARTH'S SURFACE
(A)
λ/2
λ/2
S
SPACING
λ/2
λ/2
1:1
BALUN
λ/4 MATCHING STUB
(B)

FIGURE 8.8

harness must be twisted once as shown. The feedpoint impedance is around 3000 ohms, so a quarter-wavelength matching stub is needed in order to effect a reasonable match to the coaxial transmission line. Again, for multiband operation tuned feeders are advisable.

CURTAIN ARRAYS

The curtain arrays in this section are variations of the broadside arrays. The antenna shown in Figure 8.9 is a small **Sterba curtain array**, larger versions of which are often used by high-power international short-wave broadcasters. It can be fed with either 300 ohm twin-lead, or with 75 ohm coaxial cable if a 4:1 BALUN transformer is provided at the feedpoint.

The Sterba curtain antenna can be built of wire or aluminum tubing, although the wire option is probably the most popular at HF, and tubing at VHF. An advantage of these antennas is that they can be built for frequencies in the 6–7 MHz range (where wire construction is preferred), if you have enough room, and also well into the VHF region (in which case aluminum tubing construction is preferred).

The signal is bidirectional, and is broadside to the array (in and out of the page). Gain is of the order of 6 dB, although it rises with added numbers of sections.

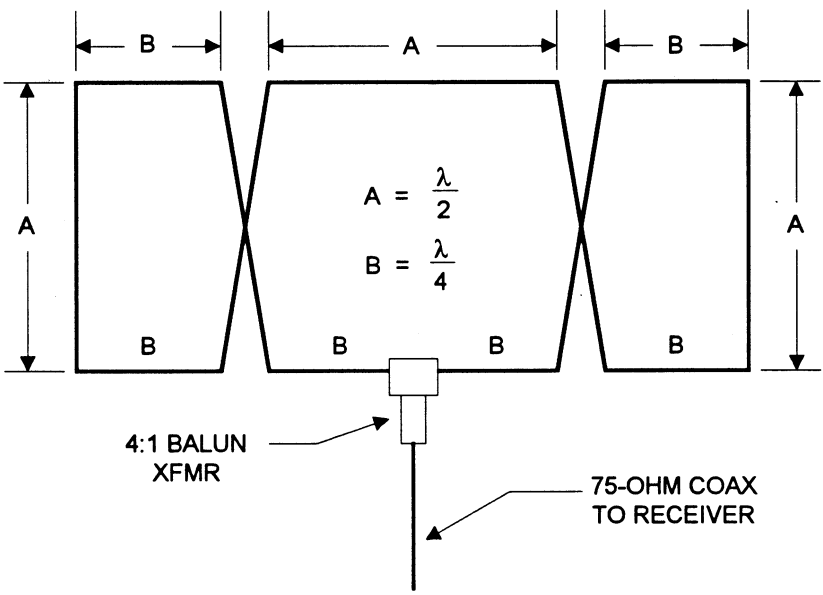


FIGURE 8.9

The antenna in Figure 8.10 (our micro-Sterba) uses elements of two different lengths ($\lambda/2$ and $\lambda/4$), labelled A and B. These lengths can be calculated from

$$A = \frac{149}{F_{\text{MHz}}} \text{ meters}$$

$$B = \frac{A}{2}$$

At 16 MHz, these lengths work out to be $A = 9.31$ m, $B = 4.66$ m; at 162 MHz, $A = 0.92$ m and $B = 0.46$ m.

The horizontal distance between horizontal elements should be 10.2–15 cm. In wire antennas, an ordinary end insulator placed between two elements will usually suffice.

The Sterba curtain concept can be extended by adding sections, which of course also increases gain. The version shown in Figure 8.10 provides gain up to 8 dB, and is basically a pair of the previous antennas back to back. The feedpoint is matched using a quarter wavelength stub and a 1:1 BALUN transformer. As with the other antennas, this is a broadside array.

The multisection Sterba curtain array shown in Figure 8.11 uses five loops, and is fed at the end with 300 ohm twin-lead (or a 4:1 BALUN connected to 75 ohm coaxial cable). Gain will be up to 9 dB or so, but at the cost of a lot of space (each A-section is a half wavelength long).

The **Bruce array** is shown in Figure 8.12. This type of array is built using a long wire folded and fed in such a manner that the current nodes (i.e. points of high current) are in the centers of the vertical elements, while the

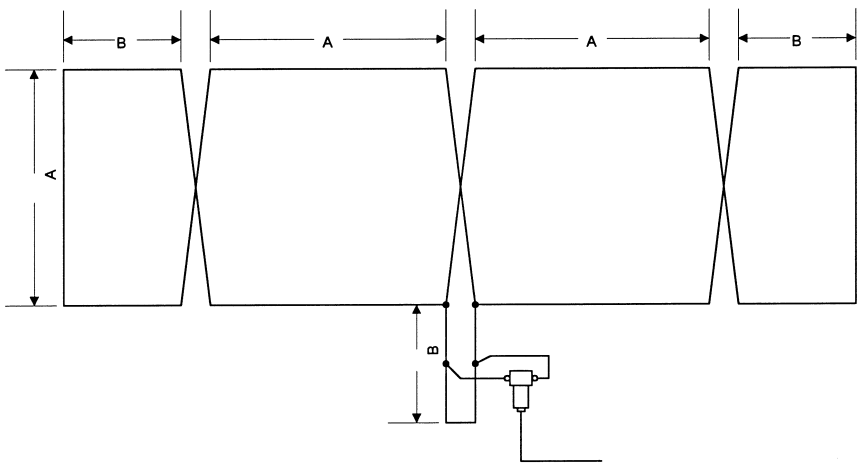


FIGURE 8.10

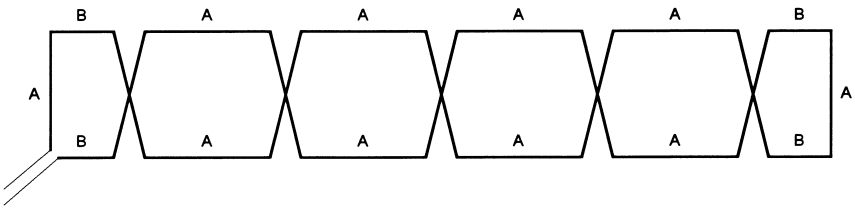


FIGURE 8.11

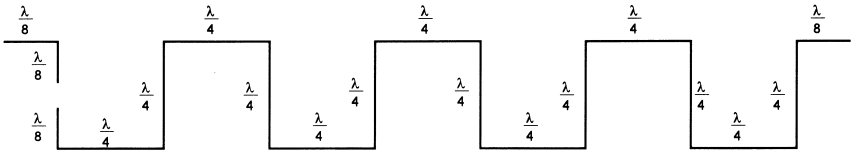


FIGURE 8.12

voltage nodes (points of low current) are in the centers of the horizontal elements. About 9 dB can be achieved with this antenna.

A variant Bruce array is shown in Figure 8.13. This antenna consists of two cross-connected sections, each of which are made of quarter-wavelength elements. The feedline is connected at the points marked X1–X2. The impedance at this point is around 700–800 ohms. If a 9:1 BALUN transformer is used, then this point can be fed with 75 ohm coaxial cable with only a small voltage standing wave ratio (VSWR), assuming single-band operation. It is also possible to feed this antenna on a single band with a quarter-wavelength matching stub, along with a 1:1 BALUN transformer. If you want to operate on two harmonically related bands, however, tuned feeders and an antenna tuner are required.

'SIX-SHOOTER' ARRAY

The array shown in Figure 8.14A is sometimes called the 'six-shooter' because it consists of six half-wavelength elements. In structure it is much

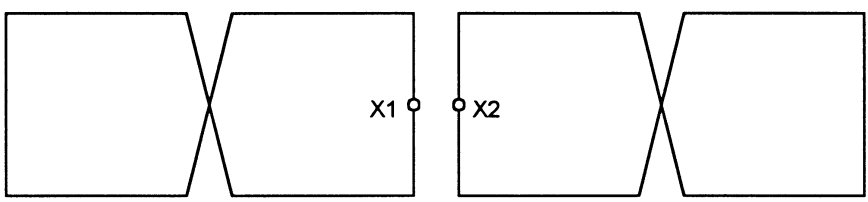
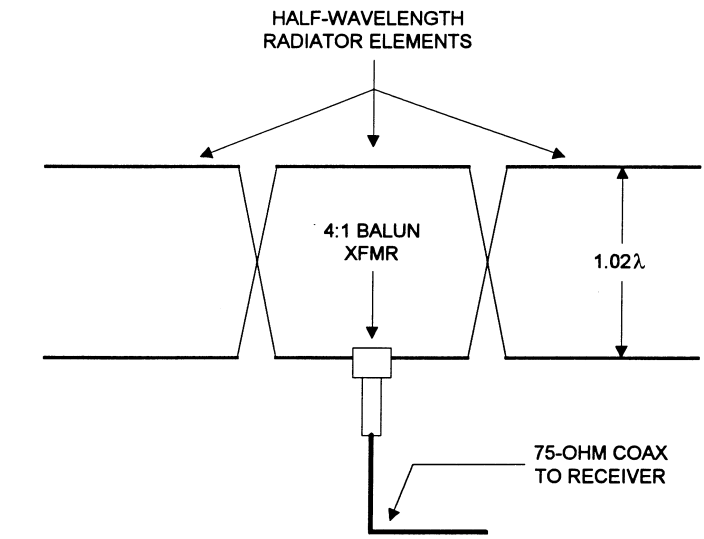


FIGURE 8.13

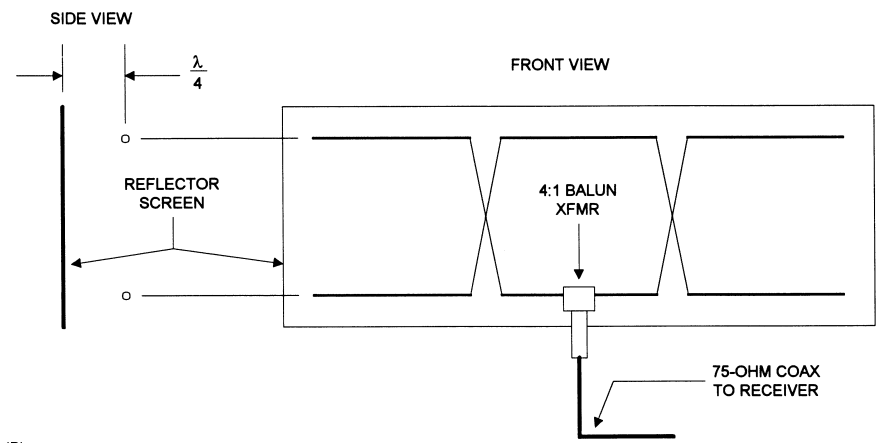
like the Sterba curtain shown earlier, with the end elements open circuited rather than connected together. This antenna is capable of gains between 7 and 8 dB.

The lengths of the elements in Figure 8.14A are calculated as shown in the figure. The horizontal elements are each calculated from

$$L_{\text{meters}} = \frac{145}{F_{\text{MHz}}} \text{ meters}$$



(A)



(B)

FIGURE 8.14

While the vertical separation between the two rows of horizontal elements is calculated from

$$L_{\text{meters}} = \frac{296}{F_{\text{MHz}}} \text{ meters}$$

These antennas are relatively easy to construct of either wire or tubing, and should be considered whenever you want gain on the cheap.

Like other broadside arrays, the 'six-shooter' is bidirectional, and sends or receives signals in and out of the page as you view the drawing. One variant that I saw on a 10 m amateur radio band 'six-shooter' is shown in Figure 8.14B. In this version, a screen is placed a quarter wavelength behind the antenna. Signals traveling toward the screen from the antenna are reflected back toward the antenna element. Because the spacing is a quarter wavelength, the signal reflected arrives back at the antenna in phase, so reinforces the signal going in the other direction. A gain improvement of 3 dB over the gain of the unscreened version is realized.

The screen can be almost any form of reflective plane. Although one could use sheets of copper or aluminum, this approach is not the best for most purposes. The reason is that the reflector then has a tremendous 'sail area' and so will be susceptible to wind problems. A better solution might be to use metal window screening (if you can still get it), or 'chicken wire' (i.e. the kind of wire used to screen chicken coops). Alternatively, one could also create an effective screen by connecting wires in a square matrix, provided that the gaps are only a small fraction of a wavelength (i.e. $\ll \lambda/8$).

THE BOBTAIL CURTAIN ARRAY

The bobtail curtain (Figure 8.15) is something of a favorite in the HF short-wave bands (with both listeners and ham operators) when very low angles of radiation are required. A low angle is the desired condition for maximizing the distance received by any given antenna (a good 'DX' antenna has an angle of radiation that barely kisses the horizon). The bobtail curtain fits this requirement rather nicely. Although it is easiest to construct for the upper HF bands, it is not beyond reason for many sites for 75/80 m band amateur radio operation or 60 m band short-wave activity.

The bobtail curtain consists of three quarter-wavelength radiators (*B*) connected at their tops by half-wavelength sections (*A*). Because the antennas are a quarter wavelength, the proper lengths are of the order of 19.82–21.34 m for the 75/80 m band and 14.63 m for the 60 m band. For the 13 m band, on the other hand, the lengths of the vertical radiators are only 2.9–3.05 m. The connecting half-wavelength sections are twice these lengths. Note that the current which is injected into the center section must split at the top, and only one-half of the line current of the center radiator flows

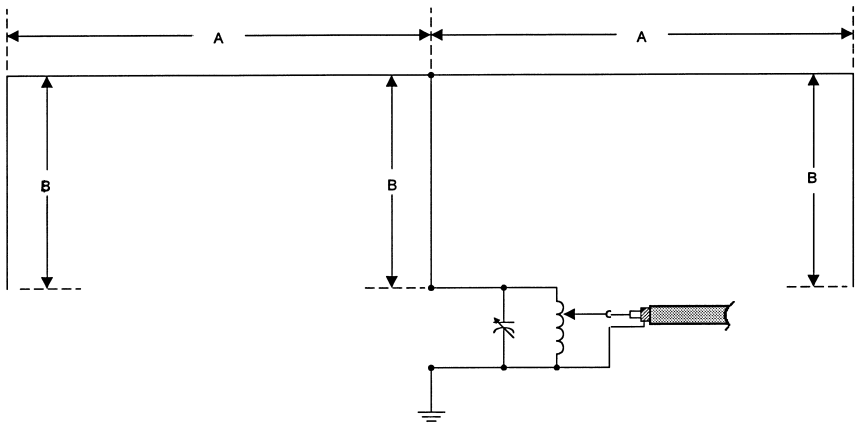


FIGURE 8.15

in each side radiator. This results in a **binomial current distribution** between the elements.

The gain of the bobtail curtain is 7–10 dB, with a figure '8' pattern broadside to the antenna. In trials some years ago, I witnessed a well-made bobtail curtain that only exhibited 5–6 dB (keeping in mind the difficulty of measuring antenna gain *in situ*), but even 5 dB is relatively decent compared to a dipole (nearly an *S*-unit on conservatively specified receivers). The directivity of the bobtail curtain is better than the dipole, but not quite as good as a set of three vertical radiators fed in-phase with each other (which will produce nearly the same pattern).

The bobtail curtain is fed at the base of the center element through a parallel tuned LC tank circuit used for impedance matching. The coaxial cable is connected to either a tap on the inductor (as shown) or to a link or primary winding (a couple of turns of wire) to the main inductor. The values of the inductor and capacitor are set such that the reactances are equal to 1100 ohms. For operation up to 20 MHz, a 50 pF variable capacitor will normally be satisfactory, although at higher frequencies the minimum capacitance of the typical 50 pF unit may be too high for the required capacitance (only a few picofarads). The inductor will vary from about 10 to 50 μH. The various inductances can be provided by either a variable inductor, or a fixed inductor with taps at the required inductance points.

THE THORNE ARRAY BOBTAIL CURTAIN ANTENNA

The Thorne Array Bobtail Curtain (TABC), shown in Figure 8.16, is basically a standard bobtail curtain array turned upside down, and fed

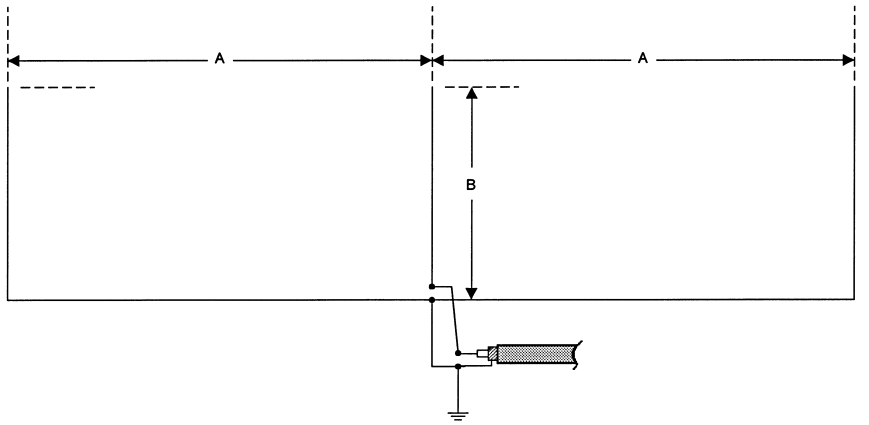


FIGURE 8.16

in a different manner. Bidirectional gains up to 5 dB have been measured on the 15 m band, although I suspect that the gain is due largely to compression of the elevation lobe of the antenna. This antenna has a very low angle of radiation, so works well for long-DX in the upper end of the short-wave spectrum. On the 15 m ham band, from a location in Texas, I have worked very loud Australian and New Zealand stations that were a lot less audible on the quarter-wavelength vertical and dipole antennas at the same location (and received much better signal reports of my own signal).

The TABC consists of three vertical quarter-wavelength ($\lambda/4$) radiators (B) separated by half-wavelength ($\lambda/2$) phasing elements (A). The two outer vertical elements are electrically connected to the corresponding phasing element, but insulated from the center vertical element. The two phasing elements are electrically connected to each other by a very short jumper wire. The center conductor of the 52 ohm coaxial cable to the receiver is connected to the base of the center vertical element, while the coaxial shield is connected to the jumper between phasing elements, and to ground.

The lengths of the two sections are $A = 91/F_{\mathrm{MHz}}$ meters and $B = 46/F_{\mathrm{MHz}}$ meters. For the 24 MHz band, the horizontal or phasing sections are 3.64 meters long while the vertical sections are 1.84 meters long.

Mechanically, the antenna is basically supported by dipole-like antenna end insulators and ropes to various masts or support structures (trees, buildings, etc.). If you prefer, the vertical sections can be replaced with 12–50 mm aluminum tubing, but you will have to provide some form of mounting for the tubing.

FOLDED DIPOLE 'X-ARRAY'

The array shown in Figure 8.17 uses four folded dipole antennas arranged in an 'X-array' pattern to provide about 6 dB gain. The antenna elements are made from 300 ohm twin-lead. The ends of the twin-lead conductors are shorted together in this type of antenna. The arrangement is two elements in each of two bays. In each bay, the inner ends of the folded dipole elements are spaced 0.2λ apart, while the two bays are spaced 1.25λ apart. The transmission line harnesses (P) are 0.45λ long. These transmission line harnesses are connected in phase with each other at the high impedance side of a 4:1 BALUN transformer; 75 ohm coaxial cable carries the signal to the receiver. All four elements in Figure 8.17 are in the same plane, and the direction of radiation/reception is perpendicular to that plane (i.e. in/out of the page).

PHASED VERTICALS

Figure 8.18 shows the basic two-element phased vertical array. This antenna uses two quarter-wavelength wire radiators erected in a vertical manner, spaced a half wavelength apart. Ropes or heavy twine can be used to support the antenna elements. The antenna element lengths are found from $L_{\text{meters}} = 75/F_{\text{MHz}}$, while the spacing between them is $D_{\text{meters}} = 150/F_{\text{MHz}}$.

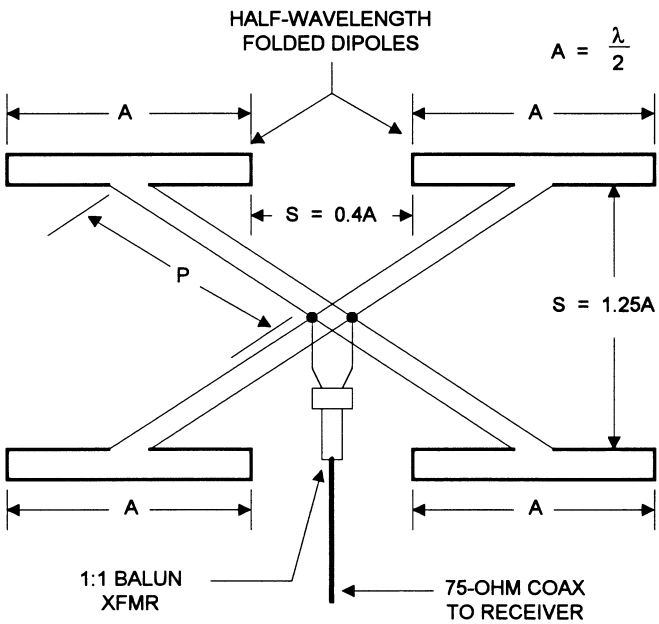


FIGURE 8.17

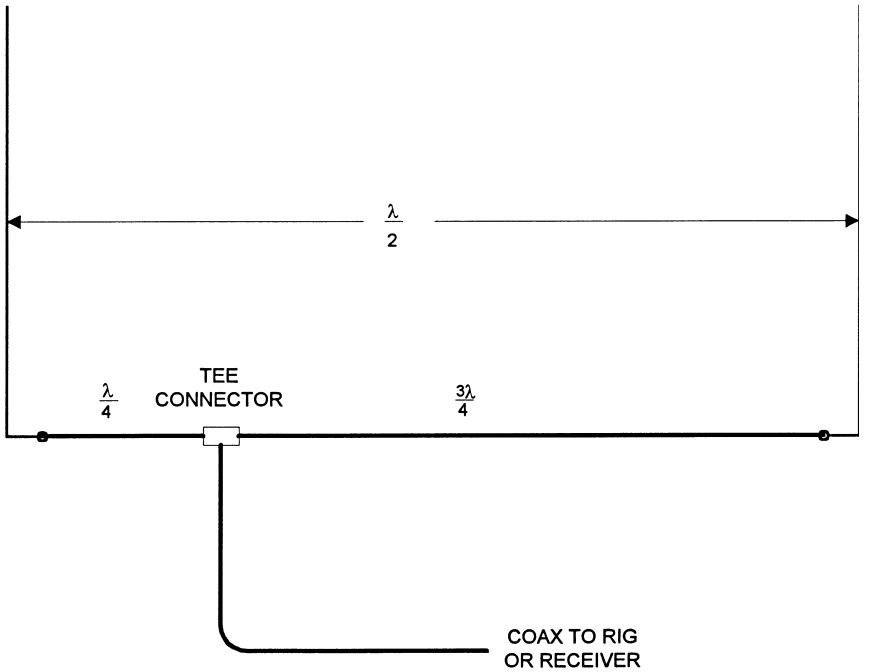


FIGURE 8.18

There are two ways to feed this antenna, and both use a coaxial cable 'tee' connector to split the signal from the rig between the two antennas. If you want to feed the antennas in phase, then the two segments of cable from the 'tee' to the antenna elements are equal lengths. But if you want to feed the elements 180° out of phase, then make one length $\lambda/4$ and the other $3\lambda/4$. In the case of in-phase feed, the directivity and gain are in and out of the page, while when the antennas are in phase the directivity and gain are bidirectional on a line between the two antennas (see again Figures 8.1E and 8.1F).

W8JK ARRAY

The W8JK array (Figure 8.19) is sometimes called the 'flat-top' beam. It is a bidirectional, end-fire array that consists of two half-wavelength in-phase collinear doublets, in parallel and fed 180° out of phase with each other, spaced between $\lambda/8$ and $\lambda/4$ apart. Each element in each collinear doublet is $\lambda/4$ long (dimension A). The view in Figure 8.20 is from above, so the plane of the dipoles is horizontal to the Earth's surface. The gain will be from 5.7 dBi at $\lambda/4$ spacing to 6.2 dBi at $\lambda/8$ spacing. Note that the spacing

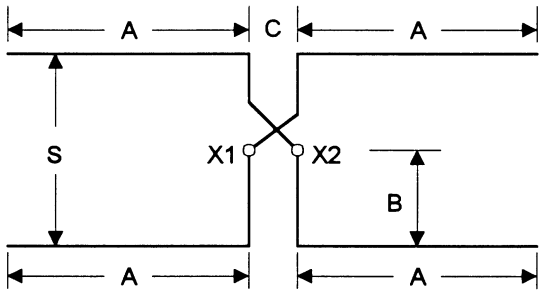


FIGURE 8.19

between the elements of the same dipole is of the order of

$$C = \frac{8.5}{F_{\text{MHz}}} \text{ meters}$$

The length of each wire element in the dipoles is found from:

$$A = \frac{71.5}{F_{\text{MHz}}} \text{ meters}$$

The feedline phasing harness between the center-feed points on the two dipoles is made of 450–600 ohm parallel transmission line (or 450 ohm twin-lead). The parallel feedline to the transmitter or receiver is attached to the phasing harness at points X1–X2, at a distance (B) of about $S/2$.

The impedance at the feedpoint (X1–X2) is several thousand ohms, so will necessarily create a rather high VSWR. An impedance matching stub can be provided at the feedpoint to improve the match to the transmission line. However, if tuned feeders are used then this antenna is capable of multiband (harmonically related) operation.

An end-fed version of the W8JK array is shown in Figure 8.20. In this case, the C-dimension is the same as calculated above, but the A-dimension

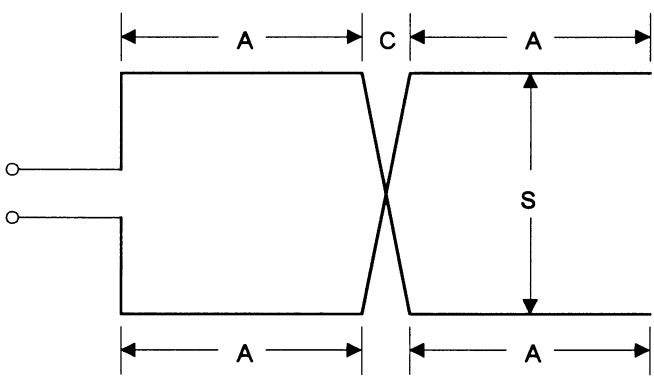


FIGURE 8.20

is not quite twice as long:

$$A = \frac{110}{F_{\text{MHz}}} \text{ meters}$$

As with the previous version, this antenna can be fed with a parallel tuned feeder, or with a matched line if an impedance matching stub is provided.

STACKED COLLINEAR ARRAYS

The collinear array antenna can be stacked as in Figure 8.21. This antenna consists of two collinear arrays (similar to Figure 8.5). The lengths of A, B and C are found from:

$$A = \frac{144}{F_{\text{MHz}}}$$

$$B = \frac{73}{F_{\text{MHz}}}$$

$$C = \frac{150}{F_{\text{MHz}}}$$

where A, B and C are in meters, and F_{MHz} is frequency in megahertz.

The two antennas are shown stacked vertically, although they can be stacked horizontally as well (although with a different pattern).

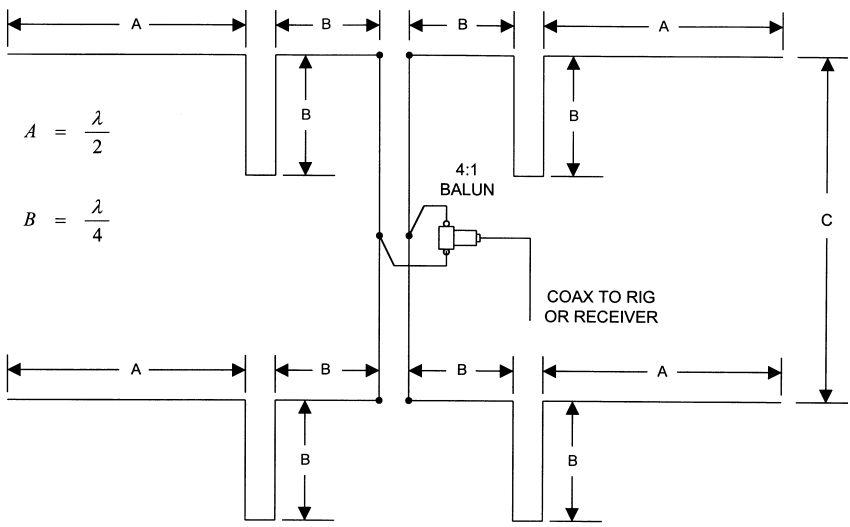


FIGURE 8.21

LARGE REFLECTOR ARRAY ANTENNA

Figure 8.22 shows a large-scale reflector array. It consists of a group of sixteen half-wavelength elements stacked vertically and horizontally. The elements are cross-connected with phasing harnesses. The spacing between elements is half wavelength, although the formula used to calculate this half wavelength is different from the half wavelength of the wire elements:

Wire elements:

$$\frac{\lambda}{2} = \frac{143}{F_{\text{MHz}}}$$

Spacing:

$$\frac{\lambda}{2} = \frac{150}{F_{\text{MHz}}}$$

The reflector screen can be made of chicken wire or any other screening material in which the holes are not more than about $\lambda/12$ in size. It is sized at least $\lambda/8$ larger than the outside dimensions of the array itself. The reflector screen is located a quarter wavelength behind the antenna array elements. The antenna array elements are made of either thin (#10 to #18) or thick wire (#000 to #10).

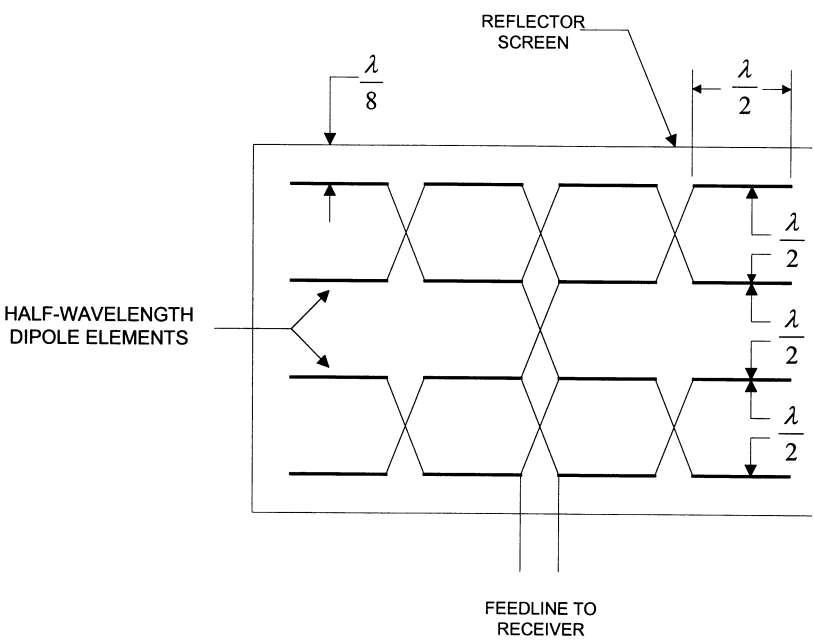


FIGURE 8.22

CHAPTER 9

Small loop antennas

In this chapter you will learn about small loop receiving antennas. The small loop antenna is almost ideal for DXing the crowded AM broadcast band and low frequency 'tropical' bands. These antennas are fundamentally different from the large loop types previously discussed and are very often the antenna of choice for low frequency work.

Large loop antennas are 0.5λ or larger and respond to the electrical field component of the electromagnetic wave. Small loop antennas, on the other hand, are $<0.1\lambda$ (some sources say 0.17λ and $<0.22\lambda$) and respond to the magnetic field component of the electromagnetic wave. One principal difference between the large loop and the small loop is found when examining the radio frequency current induced in the loop when a radio signal intersects it. In a large loop, the dimensions in each section are an appreciable portion of one wavelength, so the current will vary from one point in the conductor to another. But in a small loop, the current is the same throughout the entire loop.

The differences between small loops and large loops show up in some interesting ways, but perhaps the most striking are the directions of maximum response – the main lobes – and the directions of the nulls. Both types of loop produce figure-of-eight patterns, but in directions at right angles with respect to each other. The large loop antenna produces main lobes *orthogonal*, at right angles or 'broadside' to, the plane of the loop. Nulls are off the sides of the loop. The small loop, however, is exactly the opposite: the main lobes are off the sides of the loop (in the direction of the loop plane), and the nulls are broadside to the loop plane (see Figure 9.1A).

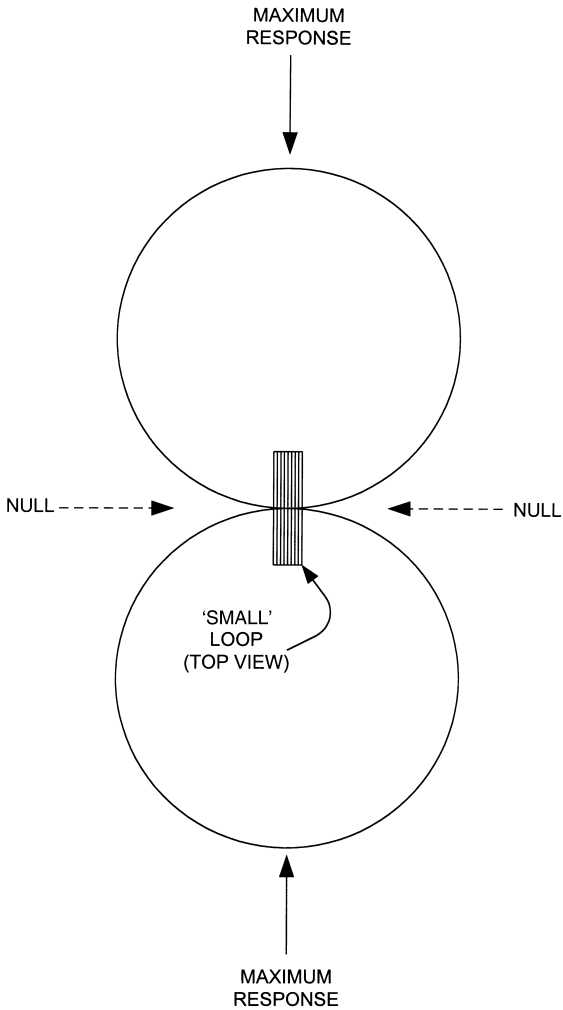


FIGURE 9.1A

Don't confuse small loop behavior with the behavior of the loopstick antenna. Loopstick antennas are made of coils of wire wound on a ferrite or powdered iron rod. The direction of maximum response for the loopstick antenna is broadside to the rod with deep nulls off the ends (Figure 9.1B). Both loopsticks and small wire loops are used for radio direction-finding and for shortwave, low frequency medium wave, AM broadcast band, and VLF DXing.

The nulls of a loop antenna are very sharp and very deep. Small changes of pointing direction can make a profound difference in the response of the antenna. If you point a loop antenna so that its null is aimed at a strong

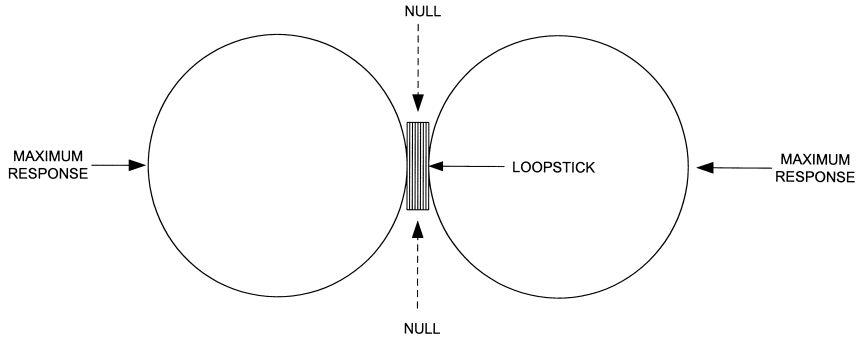


FIGURE 9.1B

station, the signal strength of the station appears to drop dramatically at the center of the notch. But turn the antenna only a few degrees one way or the other, and the signal strength increases sharply. The depth of the null can reach 10 to 15 dB on sloppy loops and 30 to 40 dB on well-built loops (20 dB is a very common value). I've seen claims of 60 dB nulls for some commercially available loop antennas. The construction and uniformity of the loop are primary factors in the sharpness and depth of the null.

One of the characteristics of the low frequency bands in which small loops are effective is the possibility of strong local interference smothering weaker ground wave and sky wave stations. As a result, you can't hear co-channel signals when one of them is very strong and the other is weak. Similarly, if a co-channel station has a signal strength that is an appreciable fraction of the desired signal, and is slightly different in frequency, then the two signals will heterodyne together and form a whistling sound in the receiver output. The frequency of the whistle is an audio tone equal to the difference in frequency between the two signals. This is often the case when trying to hear foreign broadcast band signals on frequencies (called *split frequencies*) between the standard 10 kHz spacing used in North and South America. The directional characteristics of the loop can help if the loop null is placed in the direction of the undesired signal.

Loops are used mainly in the low frequency bands even though such loops are either physically larger than high frequency loops or require more turns of wire. Loops have been used as high as VHF and are commonly used in the 10 meter ham band for such activities as hidden transmitter hunts.

Let's examine the basic theory of small loop antennas, and then take a look at some practical construction methods.

AIR CORE FRAME LOOPS ('BOX' LOOPS)

A wire loop antenna is made by winding a large coil of wire, consisting of one or more turns, on some sort of frame. The shape of the loop can be circular, square, triangular, hexagonal or octagonal. For practical reasons, the square loop seems to be most popular. With one exception, the loops considered in this section will be square so you can easily duplicate them.

The basic form of the simplest loop is shown in Figure 9.2. This loop is square, with sides the same length 'A' all around. The width of the loop ('B') is the distance from the first turn to the last turn in the loop, or the diameter of the wire if only one turn is used. The turns of the loop in Figure 9.2 are *depth wound,* meaning each turn of the loop is spaced in a slightly different parallel plane. The turns are spaced evenly across distance 'B'. Alternatively, the loop can be wound such that the turns are in the same plane (this is called *planar winding)*. In either case, the sides of the loop ('A') should be not less than five times the width ('B'). There seems to be little difference between depth and planar wound loops. The far-field patterns of the different shape loops are nearly the same if the respective cross sectional areas (πr^2 for circular loops and A^2 for square loops) are $<\lambda^2/100$.

The actual voltage across the output terminals of an untuned loop is a function of the angle of arrival of the signal *a*, as well as the strength of the signal and the design of the loop. The voltage V_o is given by:

$$V_o = \frac{2\pi ANE_f \cos(\alpha)}{\lambda}$$

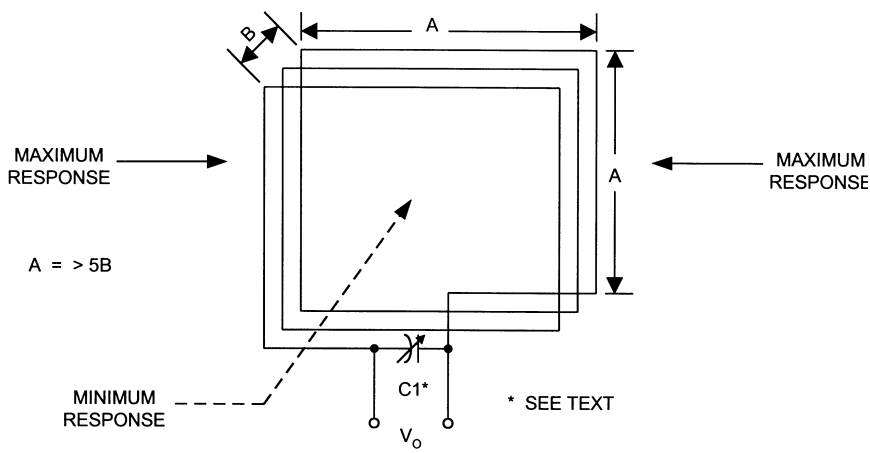


FIGURE 9.2

where:

V_o is the output voltage of the loop
A is the area of the loop in square meters (m^2)
N is the number of turns of wire in the loop
E_f is the strength of the signal in volts per meter (V/m)
a is the angle of arrival of the signal
λ is the wavelength of the arriving signal

Loops are sometimes specified in terms of the *effective height* of the antenna. This number is a theoretical construct that compares the output voltage of a small loop with a vertical piece of the same kind of wire that has a height of:

$$H_{\text{eff}} = \frac{2\pi NA}{\lambda}$$

If a capacitor (such as C1 in Figure 9.2) is used to tune the loop, then the output voltage V_o will rise substantially. The output voltage found using the first equation is multiplied by the loaded Q of the tuned circuit, which can be from 10 to 100 (if the antenna is well constructed):

$$V_o = \frac{2\pi ANE_f\, Q \cos(\alpha)}{\lambda}$$

Even though the output signal voltage of tuned loops is higher than that of untuned loops, it is nonetheless low compared with other forms of antenna. As a result, a loop preamplifier is usually needed for best performance.

TRANSFORMER LOOPS

It is common practice to make a small loop antenna with two loops rather than just one. Figure 9.3 shows such a transformer loop antenna. The main loop is built exactly as discussed above: several turns of wire on a large frame, with a tuning capacitor to resonate it to the frequency of choice. The other loop is a one or two turn coupling loop. This loop is installed in very close proximity to the main loop, usually (but not necessarily) on the inside edge not more than a couple of centimeters away. The purpose of this loop is to couple signal induced from the main loop to the receiver at a more reasonable impedance match.

The coupling loop is usually untuned, but in some designs a tuning capacitor (C2) is placed in series with the coupling loop. Because there are many fewer turns on the coupling loop than the main loop, its inductance is considerably smaller. As a result, the capacitance to resonate is usually much larger. In several loop antennas constructed for purposes of researching this chapter, I found that a 15-turn main loop resonated in the

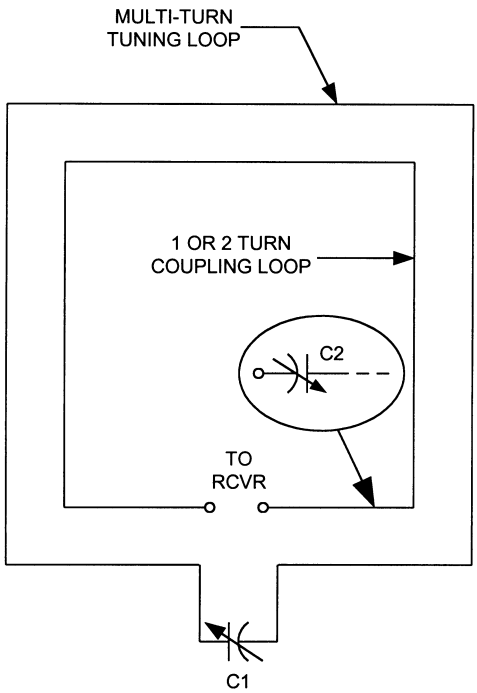


FIGURE 9.3

AM broadcast band with a standard 365 pF capacitor, but the two turn coupling loop required three sections of a ganged 3×365 pF capacitor connected in parallel to resonate at the same frequencies.

In several experiments, I used computer ribbon cable to make the loop turns. That type of cable consists of anywhere from eight to 64 parallel insulated conductors arranged in a flat ribbon shape. Properly interconnected (of which more later), the conductors of the ribbon cable form a continuous loop. It is no problem to take the outermost one or two conductors on one side of the wire array and use it for a coupling loop.

TUNING SCHEMES FOR LOOP ANTENNAS

Loop performance is greatly enhanced by tuning the inductance of the loop to the desired frequency. The bandwidth of the loop is reduced, which reduces front-end overload. Tuning also increases the signal level available to the receiver by a factor of 10 to 100 times. Although tuning can be a bother if the loop is installed remotely from the receiver, the benefits are well worth it in most cases.

There are several different schemes available for tuning, and these are detailed in Figure 9.4. The parallel tuning scheme, which is by far the most popular, is shown in Figure 9.4A. In this type of circuit, the capacitor (C1) is connected in parallel with the inductor, which in this case is the loop. Parallel resonant circuits have a very high impedance to signals on their

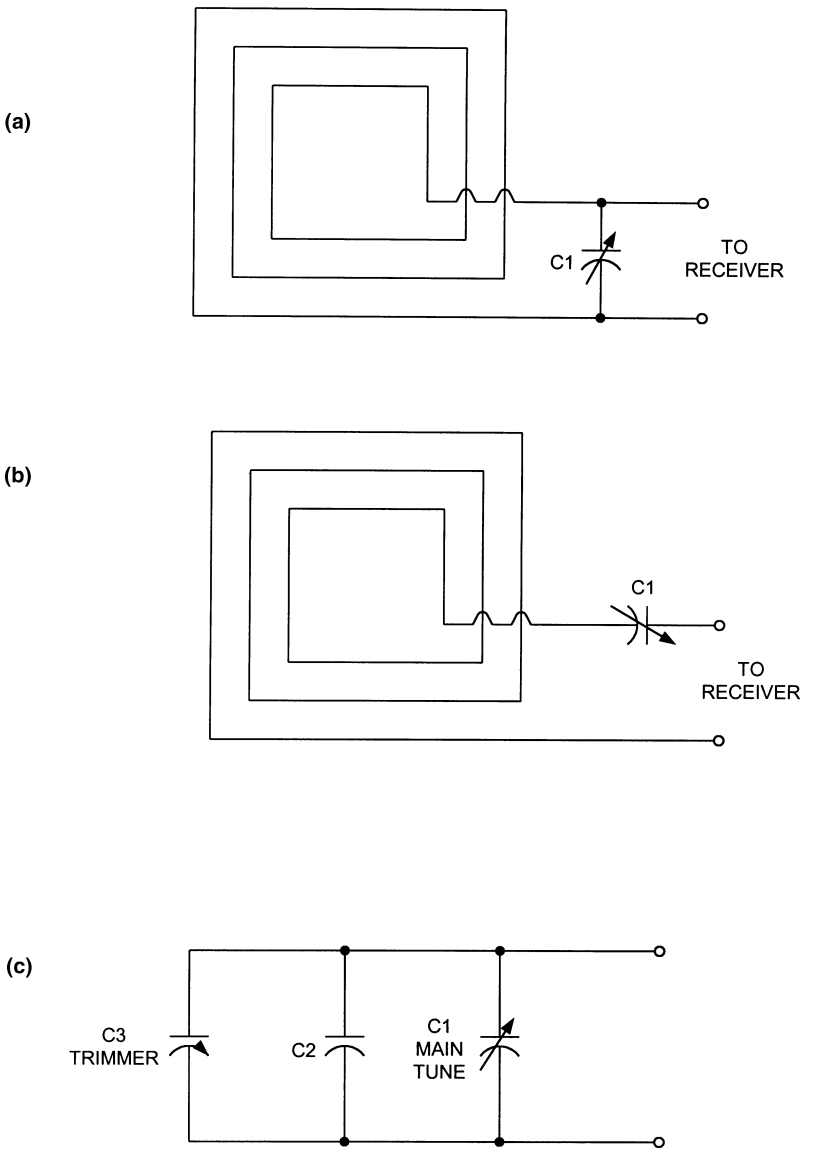


FIGURE 9.4

resonant frequency, and a very low impedance to other frequencies. As a result, the voltage level of resonant signals is very much larger than the voltage level of off-frequency signals.

The series resonant scheme is shown in Figure 9.4B. In this circuit, the loop is connected in series with the capacitor. A property of series resonant circuits is that it offers a high impedance to all frequencies except the resonant frequency (exactly the opposite of the case of parallel resonant circuits). As a result, current from the signal will pass through the series resonant circuit at the resonant frequency, but off-frequency signals are blocked by the high impedance.

There is a wide margin for error in the inductance of loop antennas, and even the precise-looking equations to determine the required values of capacitance and inductance for proper tuning are actually only estimations. The exact geometry of the loop 'as built' determines the actual inductance in each particular case. As a result, it is often the case that the tuning provided by the capacitor is not as exact as desired, so some form of compensation is needed. In some cases, the capacitance required for resonance is not easily available in a standard variable capacitor and some means must be provided for changing the capacitance. Figure 9.4C shows how this is done. The main tuning capacitor can be connected in either series or parallel with other capacitors to change the value. If the capacitors are connected in parallel, then the total capacitance is increased (all capacitances are added together). But if the extra capacitor is connected in series then the total capacitance is reduced. The extra capacitors can be switched in and out of a circuit to change frequency bands.

Tuning of a remote loop can be a bother if done by hand, so some means must be found to do it from the receiver location (unless you enjoy climbing into the attic or onto the roof). Traditional means of tuning called for using a low speed DC motor, or stepper motor, to turn the tuning capacitor. A very popular combination was the little 1 to 12 RPM motors used to drive rotating displays in retail store show windows. But that approach is not really needed today. We can use varactor voltage variable capacitance diodes to tune the circuit.

A varactor works because the junction capacitance of the diode is a function of the applied reverse bias voltage. A high voltage (such as 30 volts) drops the capacitance while a low voltage increases it. Varactors are available with maximum capacitances of 22, 33, 60, 100, and 400 pF. The latter are of most interest to us because they have the same range as the tuning capacitors normally used with loops. Look for service shop replacement diodes intended for use in AM broadcast band radios. A good selection, which I have used, is the NTE-618 device. It produces a high capacitance >400 pF, and a low of only a few picofarads over a range of 0 to 15 volts.

Figure 9.5 shows how a remote tuning scheme can work with loop antennas. The tuning capacitor is a combination of a varactor diode and two optional capacitors: a fixed capacitor (C1) and a trimmer (C2). The DC tuning voltage (V_t is provided from the receiver end from a fixed DC power supply ($V+$). A potentiometer (R1) is used to set the voltage to the varactor, hence also to tune the loop. A DC blocking capacitor (C3) keeps the DC tuning voltage from being shorted out by the receiver input circuitry.

THE SQUARE HOBBY BOARD LOOP

A very common way to build a square loop antenna is to take two pieces of thin lumber, place them in a cross shape, and then wind the wire around the ends of the wooden arms. This type of antenna is shown in Figure 9.6. The

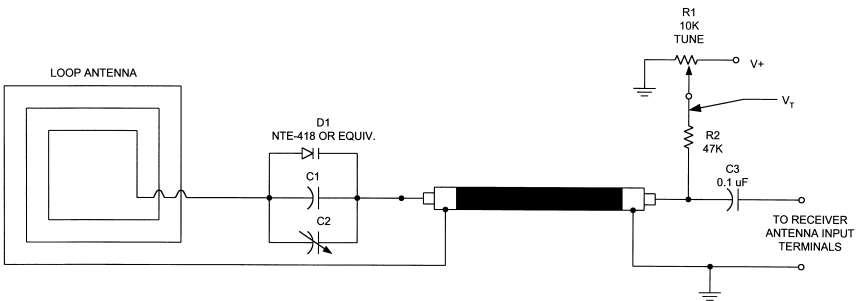


FIGURE 9.5

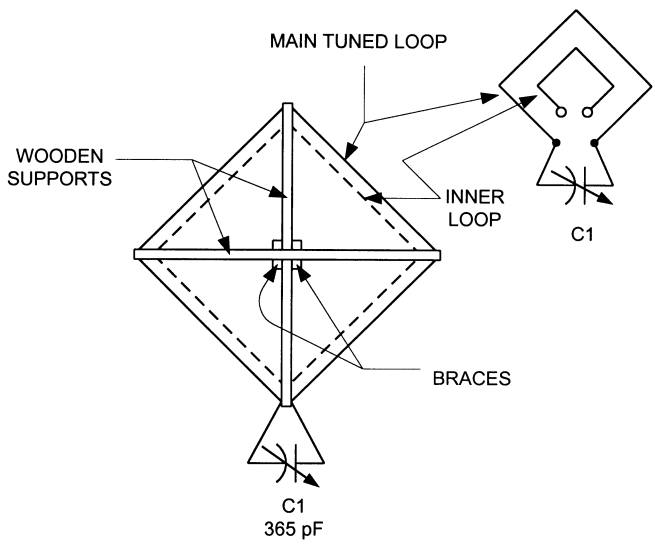


FIGURE 9.6

wooden supports can be made of $1'' \times 2''$ lumber, or some other stock. A test loop made while researching this book was made with two $3/16'' \times 3'' \times 24''$ Bass wood 'hobby board' stock acquired from a local hobby shop. Model builders use this wood as a stronger alternative to balsa wood.

The electrical circuit of the hobby board loop shown in Figure 9.6 is a transformer loop design. The main tuned loop is on the outside and consists of ten turns of #26 enameled wire. It is tuned by a 365 pF capacitor. The inner loop is used for coupling to the receiver and consists of a single turn of #22 insulated solid hook-up wire.

Details for the boards are shown in Figure 9.7. Each board is 24 inches (61 cm) long. At the mid-point (12″ or 30.5 cm), there is a $\frac{1}{4}$ inch (0.635 cm) wide, 1.5 inch (3.8 cm) long slot cut. These slots are used to mate the two boards together.

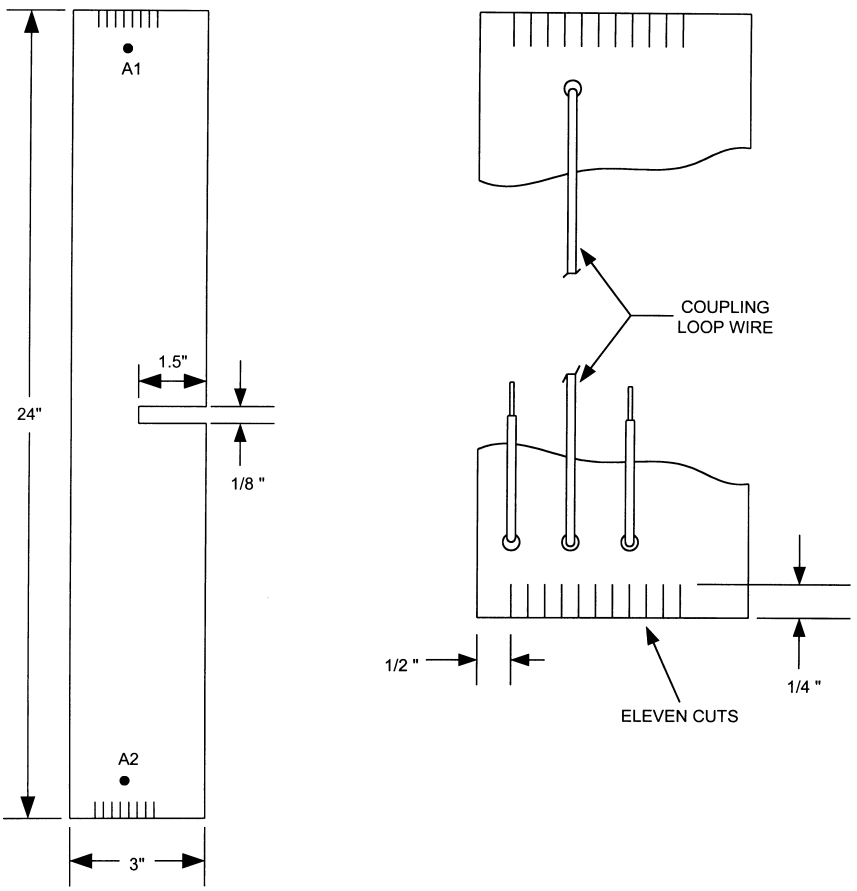


FIGURE 9.7

At each end there are ten tiny slits made by a jeweler's saw (also called a 'jig saw' in hobbyist circles) with a thin blade. These slits are just wide enough to allow a single #26 wire to be inserted without slipping. The slits are about $\frac{1}{4}$ inch (0.635 cm) long, and are $\frac{1}{4}$ inch (0.635 cm) apart. There are ten slits on both ends of the horizontal piece while the vertical piece has ten slits on the top end and 11 slits on the bottom end. The reason for offsetting the wire slits is to allow room on the other side of the 3 inch width of the vertical member for a mounting stick.

When assembling the antenna, use wood glue on the mating surfaces, square them to be at right angles to each other, and clamp the two pieces in a vice or with C-clamps for 30 minutes (or longer if the glue maker specifies). Next, glue the support blocks into place and clamp them for a similar period.

THE SPORTS FAN'S LOOP

OK, sports fans, what do you do when the best game of the week is broadcast only on a low-powered AM station – and you live at the outer edge of their service area where the signal strength leaves much to be desired? You use the sports fan's loop antenna, that's what! I first learned of this antenna from a friend of mine, a professional broadcast engineer, who worked at a religious radio station that had a pipsqueak signal but lots of fans. It really works – *one might say it's a miracle*.

The basic idea is to build a 16-turn, 24 inch (61 cm) square tuned loop (Figure 9.8) and then place the AM portable radio at the center (Figure 9.9) so that its loopstick is aimed so that its null end is broadside of the loop. When you do so, the nulls of both the loop and the loopstick are in the same direction. The signal will be picked up by the loop and then coupled to the radio's loopstick antenna. Sixteen-conductor ribbon cable can be used for making the loop. For an extra touch of class, place the antenna and radio assembly on a dining room table 'Lazy Susan' to make rotation easier. A 365 pF tuning capacitor is used to resonate the loop. If you listen to only one station, then this capacitor can be a trimmer type.

SHIELDED LOOP ANTENNAS

The loop antennas discussed thus far in this chapter have all been unshielded types. Unshielded loops work well under most circumstances, but in some cases their pattern is distorted by interaction with the ground and nearby structures (trees, buildings, etc.). In my own tests, trips to a nearby field proved necessary to measure the depth of the null because of

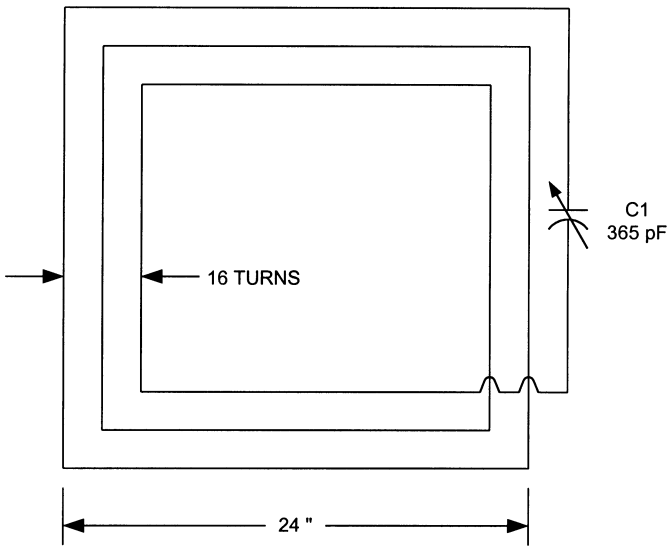


FIGURE 9.8

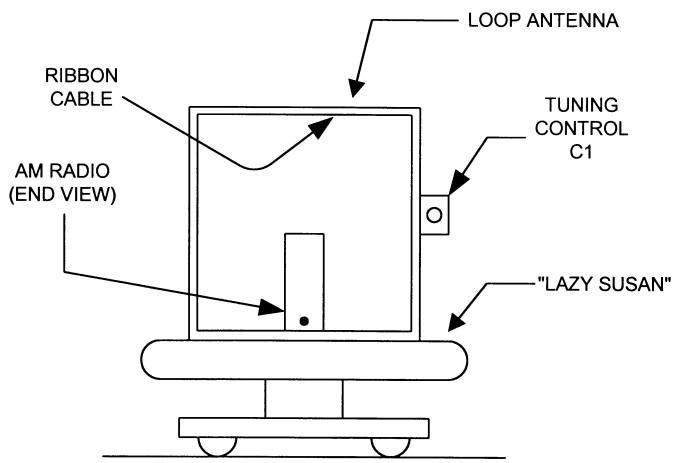


FIGURE 9.9

interaction with the aluminum siding on my house. Figure 9.10 shows two situations. In Figure 9.10A, we see the pattern of the normal 'free-space' loop, i.e., a perfect figure-of-eight pattern. But when the loop interacts with the nearby environment, the pattern distorts. In Figure 9.10B we see some filling of the notch for a moderately distorted pattern. Some interactions are so severe that the pattern is distorted beyond all recognition.

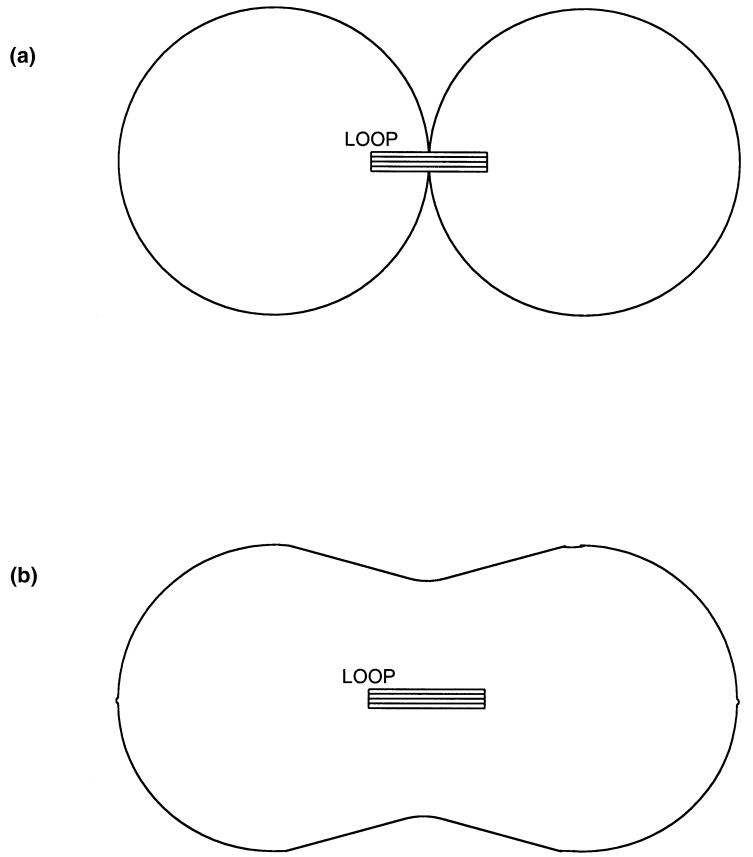


FIGURE 9.10

The solution to the problem is to reduce interaction by shielding the loop, as in Figure 9.11. Loop antennas operate on the magnetic component of the electromagnetic wave, so the loop can be shielded against voltage signals and electrostatic interactions. In order to prevent harming the ability to pick up the magnetic field, a gap is left in the shield at one point.

There are several ways to shield a loop. You can, for example, wrap the loop in adhesive-backed copper foil tape. Alternatively, you can wrap the loop in aluminum foil and hold it together with tape. Another method is to insert the loop inside a copper or aluminum tubing frame. The list seems endless. Perhaps one of the most popular methods is to use coaxial cable to make a large single turn loop. Figure 9.12 shows this type of loop made with RG-8/U or RG-11/U coaxial cable. The cable is normally supported by wooden cross arms, as in the other forms of loop, but they are not shown

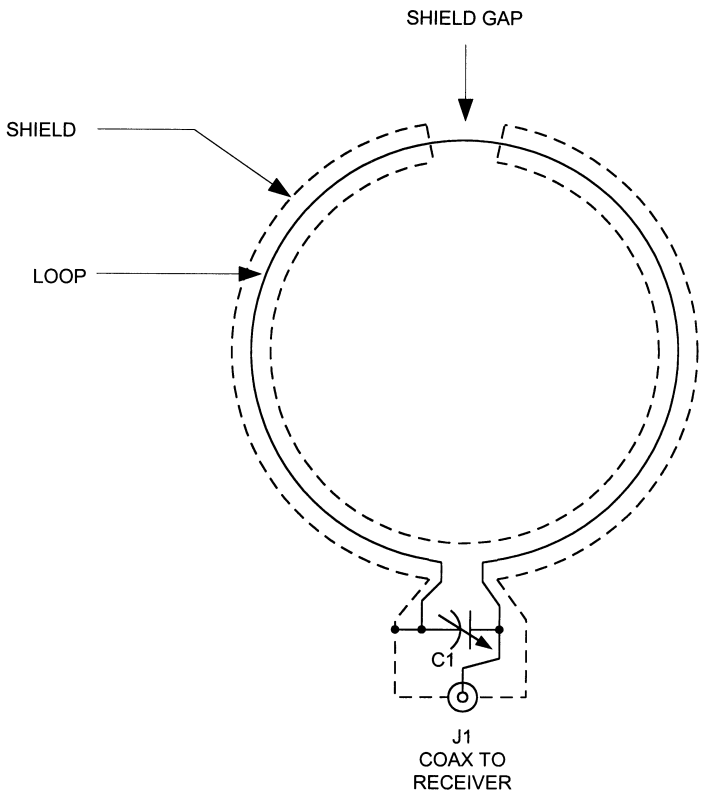


FIGURE 9.11

here for sake of simplicity. Note that, at the upper end, the coaxial cable shields are not connected.

Another example is the antenna shown in Figure 9.13. This antenna is made of wide metal conductors. Examples include the same type of hobbyist's brass stock as used above. It can also be copper foil or some other stock that can be soldered. Some electronic parts stores sell adhesive backed foil stock used for making printed circuit boards. The foil can be glued to some flat insulating surface. Although K-inch plywood springs to mind immediately, another alternative is found in artists' supplies stores. Ordinary poster board is too floppy to stand up, but poster board glued to a Styrofoam backing can be used. It is extremely easy to work with using X-acto knives and other common household tools.

Two controls are used on this antenna. Capacitor C1 tunes the loop to the resonant frequency of the desired station. Potentiometer R1 is used as a phasing control. The dimensions of the antenna are not terribly critical,

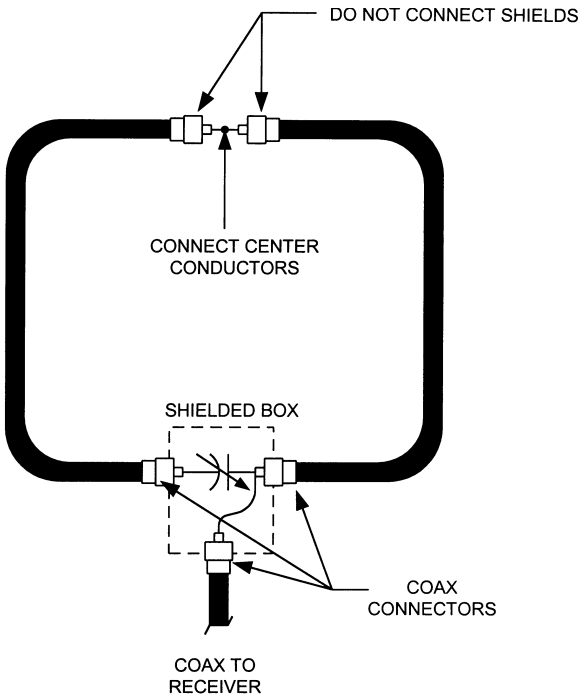


FIGURE 9.12

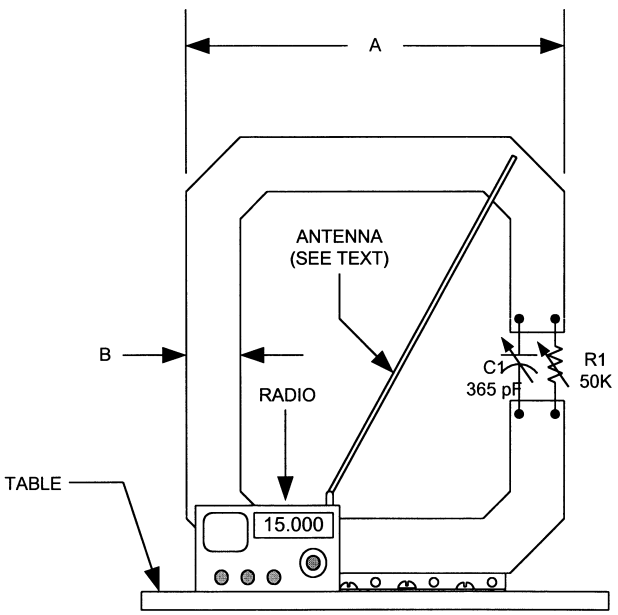


FIGURE 9.13

although some guidelines are in order. In the Villard article, he recommended a 40 cm (15.75-inch) square loop ('A'). If the loop is 7.62 cm (three inches) wide, the antenna will resonate at 15 MHz with around 33 pF of capacitance. If the dimensions are increased to A = 91 cm (36 inches) and B = 10.16 cm (four inches), then the inductance increases and only 28 pF are needed at 15 MHz. The larger size loop can be used at lower frequencies as well. For example, the 91 cm loop will resonate at 6 MHz with 177 pF.

To use this antenna, position the radio's telescopic antenna close and adjacent to the loop *but not touching it*. The loop antenna can be rotated to find the best position to either null or enhance a particular station. The 'Lazy Susan' idea will work well in this case.

TESTING YOUR LOOP ANTENNA

When each loop prototype was completed, I tested it on the AM broadcast band over several evenings. The same procedure can be used with any loop. A strong local signal at 1310 kHz served to check the pattern. The station and my home were located on US Geological Survey 7.5 minute quadrangle maps of my area (or the equivalent Ordnance Survey maps in the UK). The maps had adjacent coverage, so the compass bearing from my location to the station could be determined. Checking the antenna showed an S7/S8 signal when the loop was endwise to the station – that is, the station was in one of its lobes. Rotating the loop so that its broadside faced the direction of the station dropped the signal strength to less than S1, and frequently bottomed out the meter. Because my receiver has a 3 dB/S-unit calibration on the S-meter, I figured the null to be more than 20 dB, although it will take a bit more experimentation to find the actual depth. This test is best done during daylight hours, I found out, because there is always a residual sky wave cacophony on the AM band that raises the S-meter 'floor' in the null.

USING A LOOP ANTENNA

Most readers will use a loop for DXing rather than hidden transmitter hunting, navigation, or other RDF purposes. For the DXer, there are actually two uses for the loop. One is when you are a renter or live in a community that has routine covenants against outdoor antennas. In this situation, the loop will serve as an active antenna for receiving AM broadcast band and other low frequency signals without the neighbors or landlord becoming annoyed.

The other use is illustrated by the case of a friend of mine. He regularly tunes in clear channel WSM (650 kHz, Nashville) in the wee hours between

Saturday evening ('Grand Ole Opry' time) and dawn. However, that 'clear' channel of WSM isn't really so clear, especially without a narrow filter in the receiver. He uses a loop antenna to null out a nearby 630 kHz signal that made listening a bit dicey, and can now tape his 1940s/1950s vintage country music.

It isn't necessary to place the desired station directly in the main lobes off the ends of the antenna, but rather place the nulls (broadside) in the direction of the offending station that you want to eliminate. So what happens if the offending station and the desired station are in a direct line with each other with your receiving location in the middle between them? Both nulls and lobes on a loop antenna are bidirectional, so a null on the offending station will also null the desired station in the opposite direction.

One method is to use a sense antenna to spoil the pattern of the loop to a cardioid shape. Another method is to use a spoiler loop to null the undesired signal. The spoiler loop is a large box loop placed one to three feet (found experimentally) behind the reception loop in the direction of the offending signal. This method is detailed in Figure 9.14. The small loopstick may be the antenna inside the receiver, while the large loop is a box loop such as the sports fan's loop. The large box loop is placed about one to three feet behind the loopstick and in the direction of the offending station. The angle with respect to the line of centers should be 60° to 90°, which is also found experimentally. It's also possible to use two air core loops to produce an asymmetrical receiving pattern.

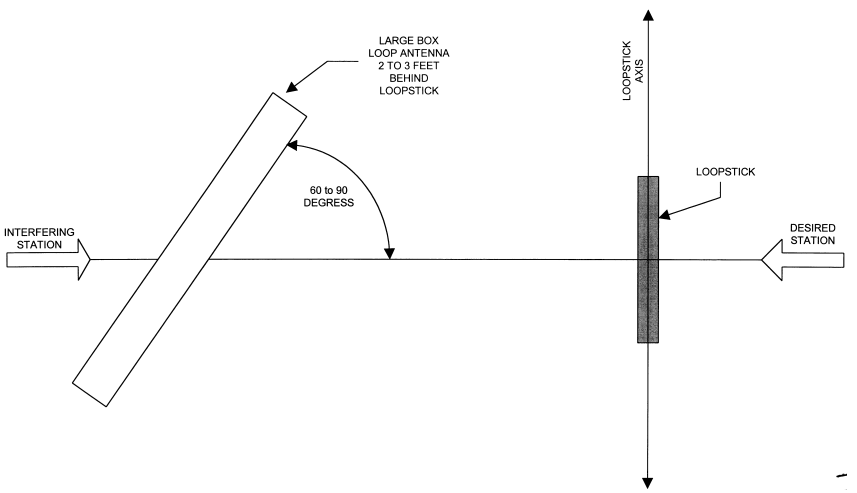


FIGURE 9.14

LOOP PREAMPLIFIERS

All small loop antennas produce a weak output signal, so a loop preamplifier is indicated for all but the most sensitive receivers. The preamplifier can be mounted either at the receiver or the antenna, but it is most effective when mounted at the antenna (unless the coax to the receiver is short).

SHARPENING THE LOOP

Many years ago the *Q-multiplier* was a popular add-on accessory for a communications receiver. These devices were sold as Heathkits and many construction projects were seen in magazines and amateur radio books. The Q-multiplier has the effect of seeming to greatly increase the sensitivity of a receiver, as well as greatly reducing the bandwidth of the front end. Thus, it allows better reception of some stations because of increased sensitivity and narrowed bandwidth.

A Q-multiplier is an active electronic circuit placed at the antenna input of a receiver. It is essentially an Armstrong oscillator, as shown in Figure 9.15, that doesn't quite oscillate. These circuits have a tuned circuit (L1/C1) at the input of an amplifier stage, and a feedback coupling loop (L3). The

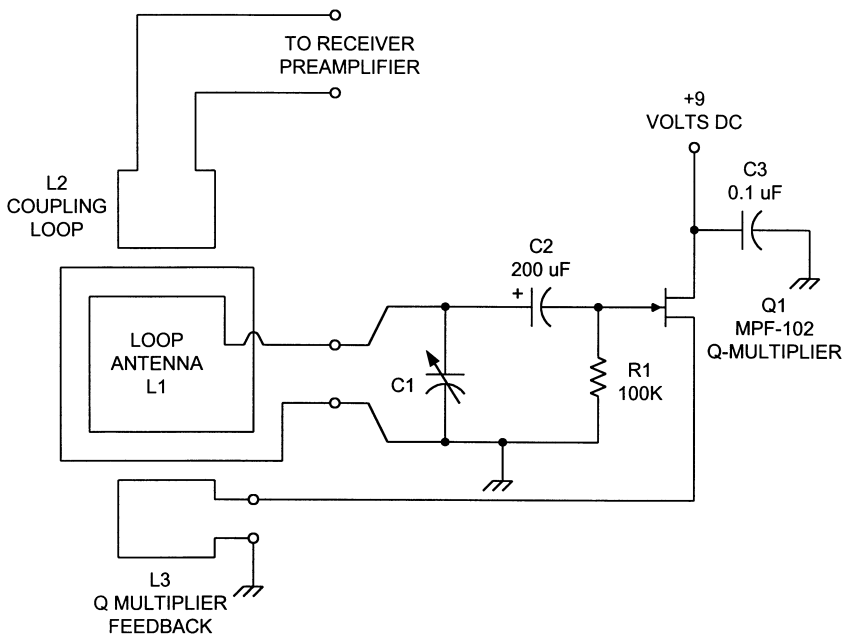


FIGURE 9.15

degree of feedback is controlled by the coupling between L1 and L3. The coupling is varied both by varying how close the two coils are, and their relative orientation with respect to each other. Certain other circuits use a series potentiometer in the L3 side that controls the amount of feedback.

The Q-multiplier is adjusted to the point that the circuit is just on the verge of oscillating, but not quite. As the feedback is backed away from the threshold of oscillation, but not too far, the narrowing of bandwidth occurs as does the increase in sensitivity. It takes some skill to operate a Q-multiplier, but it is easy to use once you get the hang of it and is a terrific accessory for any loop antenna.

CHAPTER 10

Yagi beam antennas

The Yagi beam antenna (more correctly, the Yagi–Uda antenna, after both of the designers of Tohoku University in Japan 1926) is *unidirectional*. It can be vertically polarized or horizontally polarized with little difference in performance (other than the polarization!). The Yagi antenna can be rotated into position with little effort. Yet the Yagi antenna shows power gain (so it puts out and receives a stronger signal), reduces the interfering signals from other directions, and is relatively compact.

COMPOSITION OF A BEAM ANTENNA

The Yagi antenna is characterized by a single driven element which takes power from the transmitter (or is connected to the receiver), plus one or more *parasitic elements*. The parasitic elements are not connected to the driven element, but rather receive their power from the driven element by indirect means. The indirect means is that they intercept the signal, and then re-radiate them.

The minimalist two element beam antenna may be composed of either a driven element and a reflector, or a driven element and a director. The reflector and directors are known as parasitic elements.

The *parasitic reflector* is three to five per cent *longer* than the half wavelength driven element. It provides power gain in the direction away from itself. It is inductive in reactance and lagging in phase.

The *parasitic director* is three to five per cent *shorter* than the half wavelength driven element. It provides power gain in its own direction. It is capacitive in reactance and leading in phase.

The factors that affect the phase difference between the direct and re-radiated signals is determined principally by the *element length* and the *spacing* between the elements. Proper adjustment of these factors determines the gain and the front-to-back ratio that is available.

The presence of a parasitic element in conjunction with a driven element tends to reduce the feedpoint impedance of the driven element for close spacings ($<\lambda/2$) and increase it for greater than $\lambda/2$ spacing. In general, beam antennas have element spacing of 0.1λ to 0.25λ (with 0.15λ to 0.19λ being common), so the impedance will be lower than the nominal impedance for a half wavelength dipole.

TWO-ELEMENT YAGI ARRAY

Figure 10.1 shows the two-element Yagi array antenna. This particular one uses a driven element and a director, so the direction of maximum signal is in the direction of the director. The gain of a two-element Yagi is about 5.5 dBd (gain above a dipole) for spacing less than 0.1λ and the parasitic element is a director. For the case where a reflector is used, the gain peak is 4.7 dBd at about 0.2λ spacing.

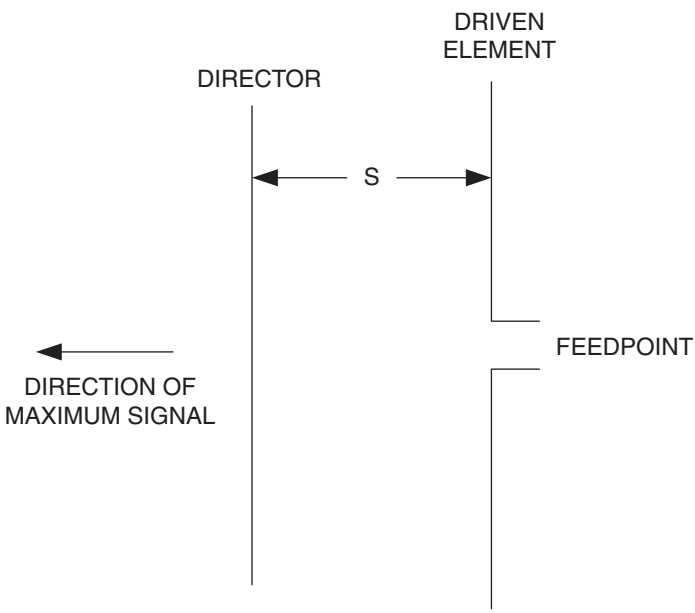


FIGURE 10.1

The difference between reflector and director usage is quite profound. The usual curve shows the director with higher gain, but it is more responsive to element spacing. The reflector has less gain, but is more tolerant of spacing errors.

The front-to-back ratio of the beam antenna is poor for two-element antennas. A compromise spacing of 0.15λ provides front-to-back ratios of 5 to 12 dB.

The feedpoint resistance of the antenna is clearly not 73 ohms as would be implied by the use of a half wavelength dipole for a driven element. The feedpoint impedance will vary roughly linearly from about 5 ohms at a spacing of 0.05λ to about 30 ohms for spacings of about 0.15λ. Above 0.15λ the differences between director and reflector implementations takes place. A reflector two-element beam feedpoint impedance will increase roughly linearly from 30 ohms at 0.15λ to about 45 ohms at 0.25λ spacing. The director implementation is a little less linearly related to spacing, but varies from about 30 ohms at 0.15λ to about 37 ohms at 0.25λ spacing.

Element lengths

The element lengths for a two-element Yagi beam are given below:

Director: $$\text{Director} = \frac{138.6}{F_{\text{MHz}}}$$

Driven element: $$\text{D.E.} = \frac{146}{F_{\text{MHz}}}$$

Spacing: $$\text{Spacing} = \frac{44.98}{F_{\text{MHz}}}$$

Where:

Director is the length of the director
D.E. is the length of the driven element in meters (m)
Spacing is the spacing between the elements in meters (m)
F_{MHz} is the frequency in megahertz.

These element lengths will result in 0.15λ spacing, which is considered about ideal.

THREE-ELEMENT YAGI BEAM

Figure 10.2 shows a Yagi antenna made up of a half wavelength driven element, a reflector and a director. The gain of the array and the front-to-back ratio peaks at a particular boom length (boom not shown), which is indicative of the spacing between the elements. Maximum gain occurs at a boom length of 0.45λ. An example of a three-element Yagi antenna built on a 0.3λ boom will provide 7 to 8 dBd forward gain, and a front-to-back ratio of 15 to 28 dB depending on the element tuning.

The feedpoint impedance of the three-element beam is about 18 to 25 ohms, so some means must be provided for adjusting the impedance to the 52 ohm coaxial cable.

Element lengths

Director: $$\text{Director} = \frac{140.7}{F_{\text{MHz}}}$$

Driven element: $$\text{D.E.} = \frac{145.7}{F_{\text{MHz}}}$$

Reflector: $$\text{Reflector} = \frac{150}{F_{\text{MHz}}}$$

Spacing: $$\text{Spacing} = \frac{43.29}{F_{\text{MHz}}}$$

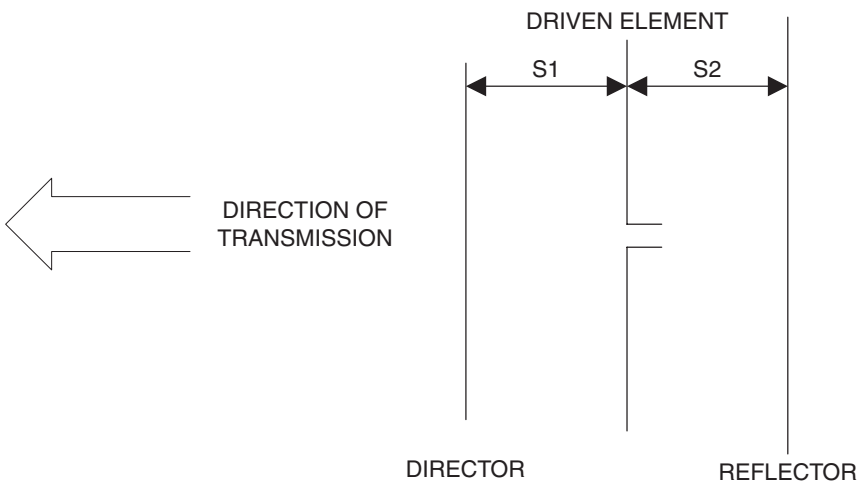


FIGURE 10.2

Where:

Director is the length of the director in meters (m)
D.E. is the length of the driven element in meters (m)
Reflector is the length of the reflector element in meters (m)
Spacing is the spacing of the elements in meters (m).

FOUR-ELEMENT YAGI ANTENNA

Figure 10.3 shows a four-element Yagi antenna. There is a tremendous increase in forward gain by adding a second director to the three-element case, but the front-to-back ratio is poorer unless the spacing is increased from 0.15λ to about 0.25λ. When all elements are spaced 0.15λ apart, the front-to-back ratio is only about 10 dB, but at 0.25λ the front-to-back ratio increases to 27 dB.

Element lengths

The dimensions calculated from the equations below will yield a forward gain of about 9.1 dBd, with a front-to-back ratio of about 27 dB.

Director: $$\text{Director} = \frac{138.93}{F_{\text{MHz}}}$$

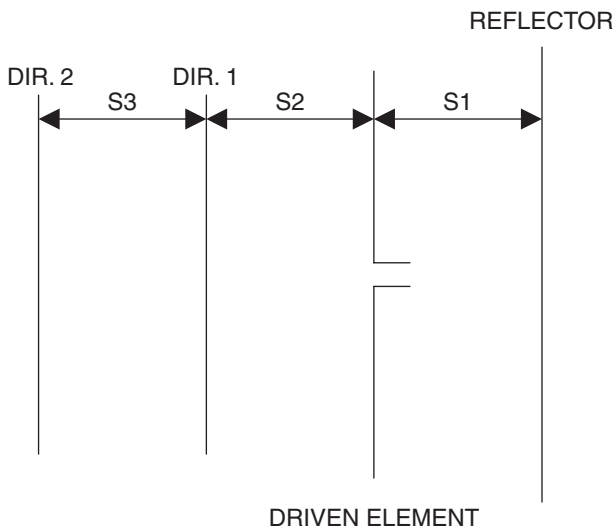


FIGURE 10.3

Driven element: $$\text{D.E.} = \frac{143.65}{F_{\text{MHz}}}$$

Spacing S1: $$\text{S1} = \frac{60.95}{F_{\text{MHz}}}$$

Spacings S2 and S3: $$\text{S2} = \text{S3} = \frac{75}{F_{\text{MHz}}}$$

Where:

Director is the length of the director element in meters (m)
D.E. is the length of the driven element in meters (m)
Reflector is the length of the reflector element in meters (m)
S1, *S2* and *S3* are in meters (m).

SIX-ELEMENT YAGI ANTENNA

Computer studies of Yagi antenna arrays demonstrate that five-element antennas are little more than four-element antennas, despite the extra director. But the addition of two additional directors adds significantly to the gain. The six-element antenna (Figure 10.4) shows a gain of nearly 10.5 dBd, and a front-to-back ratio of nearly 35 dB. Unfortunately, a six-element antenna is quite large, even with 0.15λ element spacing. At 14 MHz the antenna soars to about 16 meters boom length.

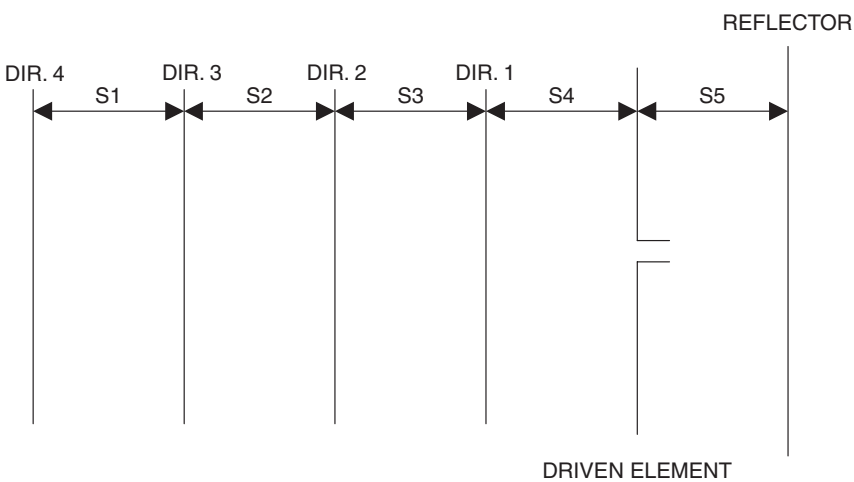


FIGURE 10.4

Element lengths

Director: $$\text{Director} = \frac{134.39}{F_{\text{MHz}}}$$

Driven element: $$\text{D.E.} = \frac{144.05}{F_{\text{MHz}}}$$

Reflector: $$\text{Reflector} = \frac{148.56}{F_{\text{MHz}}}$$

Element spacing: $$\text{Spacing} = \frac{44.8}{F_{\text{MHz}}}$$

Where:
Director is the length of the director elements in meters (m)
D.E. is the length of the driven element in meters (m)
Reflector is the length of the reflector element in meters (m)
Spacing is the spacing between the elements in meters (m).

IMPEDANCE MATCHING THE BEAM ANTENNA

The feedpoint impedance of most beam antennas is lower than the feedpoint impedance of a half wavelength dipole (72 ohms), despite the fact that the half wavelength dipole is a driven element. The feedpoint impedance may be as low as 18 to 20 ohms, and as high as 37 ohms. At 37 ohms there is a reasonable match to 52 ohm coaxial cable (1.41:1), but at 25 ohms the VSWR is more than 2:1. The typical solid-state transmitter will shut down and produce little RF power at this VSWR. What is needed is a means of matching the impedance of the beam to 52 or 75 ohm coaxial cable.

The *gamma match* is shown in Figure 10.5. It consists of a piece of coaxial cable connector such that its shield is to the center point on the radiating element (L), and its center conductor goes to the matching device. The dimensions of the gamma match of Figure 10.5 are as follows:

(L is the driven element length)

$$A = \frac{L}{10}$$

$$B = \frac{L}{70}$$

Where:
L, A and B are in meters (m).

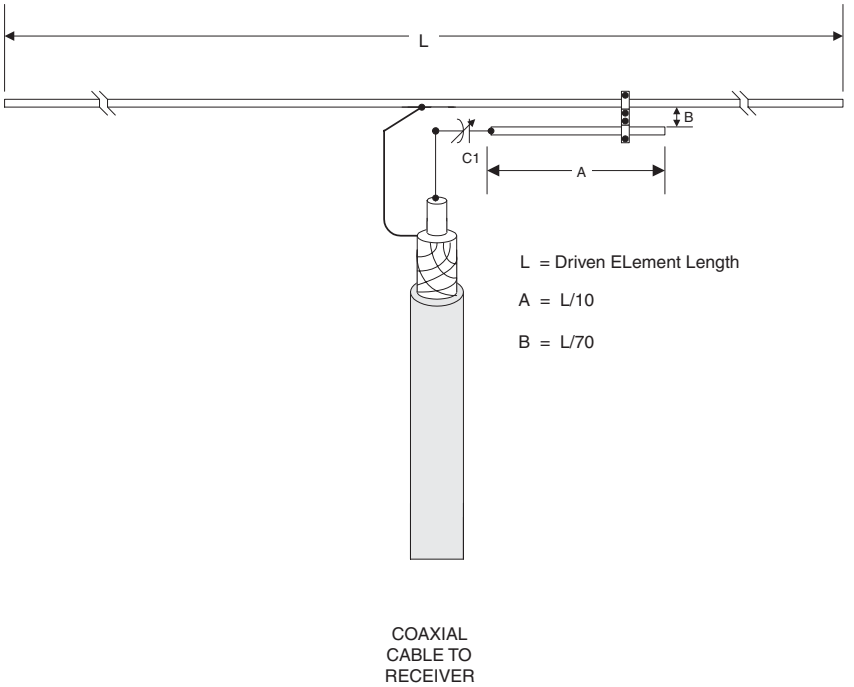

L
B
C1
A
L = Driven ELement Length
A = L/10
B = L/70
COAXIAL
CABLE TO
RECEIVER

FIGURE 10.5

CHAPTER 11

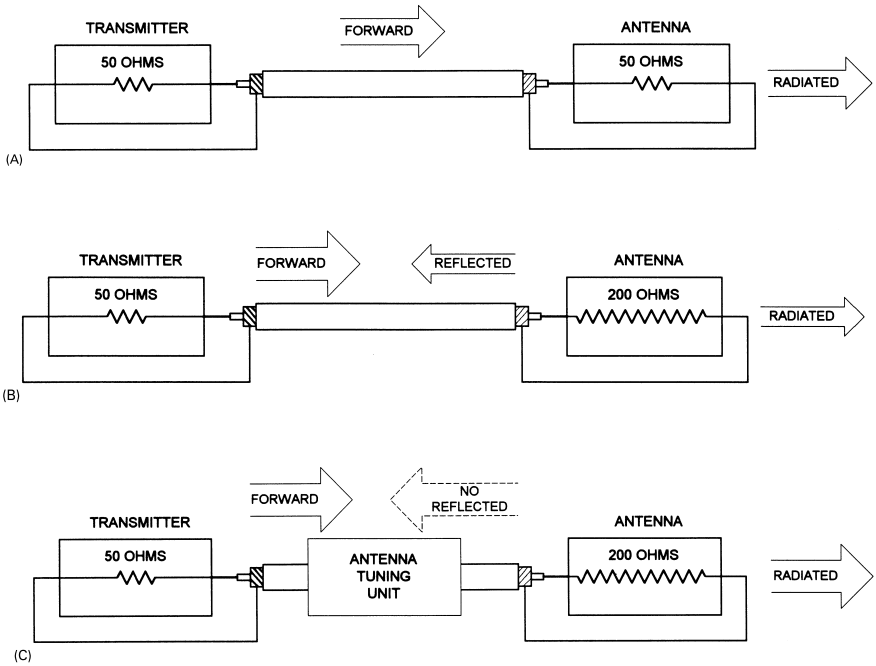
Impedance matching

It is common practice to match the impedances of the antenna, transmission line, and either the receiver or the transmitter. The reason can be seen in Figure 11.1. Although the example selected is for a transmitting antenna, because of reciprocity the same notions also apply to the receive antenna, although power flows are reversed (and considerably smaller).

The case shown in Figure 11.1A is for the matched case, i.e. the transmitter output impedance, the transmission line impedance, and the antenna feedpoint impedance (which is the 'load') are all the same (e.g. 50 ohms). The thickness of the arrow is intended to represent the relative power level.

In this situation, forward power from the transmitter is sent toward the load end of the transmission line. When the radio frequency (RF) signal reaches the load it is *completely absorbed*. Some of it will go towards heating the load, and part of it will be radiated. The loss to heating in a well-designed, properly installed antenna is negligible, so it is not unreasonable to show the arrow representing the radiated portion as the same size as the forward power. The implication of the system being matched is that all of the power from the transmitter is radiated and can become a useful signal.

When the load and source are mismatched, as in Figure 11.1B, not all of the foward power is absorbed by the load. Some of it is reflected back down the transmission line toward the transmitter. When this reflected wave interferes with the forward waves still propagating up the line, the result is **standing waves** (see Chapter 2). The measure of the standing waves is the **standing wave ratio** (SWR, or VSWR if voltage is measured).



The existence of the reflected power wave has some implications in practical systems. First, it represents a loss. The reflected power is not available to contribute to the radiated signal. If no other losses occur, then this loss might prove tolerably low, but that is rarely the case. At some frequencies and in some situations, the goal is to squeeze out as much signal as possible.

Second, some transmitters, mostly modern designs with solid-state final RF power amplifiers, are intolerant of high VSWR loads. In the early days of solid-state transmitters, the first of which were CB rigs, the expensive output transistors would often blow out when even a transient high VSWR situation occurred. Older vacuum tube ('valve' to UK readers) designs were more tolerant of a high VSWR. Although those systems experienced the VSWR loss problem, the transmitter was not harmed. Modern transmitters, on the other hand, now include **automatic load control** (ALC) or other circuitry to begin reducing RF output power as the VSWR climbs. The shut down 'knee' usually starts somewhere around a VSWR of 1.5:1, and the rig will be completely shut down at some VSWR between 2.5:1 and 3:1. One rig produced 100 W of power at a VSWR of 1:1, but only 100 mW at a VSWR of 2.5:1.

A third reason is that losses in coaxial cable can be quite severe, especially at higher frequencies. Those losses increase markedly with increasing

VSWR, especially in the lower grades of coaxial cable normally used for radio antennas.

If the antenna feedpoint impedance does not match the feedline impedance, then a VSWR proportional to their ratio is created. Unfortunately, it is not always possible to create an antenna with a feedpoint impedance equal to the characteristic impedance of the feedline. Unless some means is found to effect the match, then all of the gremlins associated with a high VSWR occur.

The solution to the problem is to provide an intermediate impedance matching device (Figure 11.1C) between the transmitter (or receiver) and the antenna. The matching device might be an inductor–capacitor (LC) network, a transformer (e.g. a BALUN), or a transmission line matching section. The impedance matching device can be considered as a 'black box' between two resistances (R1 and R2, as in Figure 11.2). The property of the box is to provide the transformation that makes each resistor think it is looking at another resistor of the same value as itself. Although these devices are bilateral, i.e. the transformation occurs in both directions, it is common practice to use R1 to represent the transmitter output impedance or the receiver antenna input impedance, and R2 to represent the load impedance (e.g. antenna feedpoint impedance). In most cases, the value of R1 in transmitters is 50 ohms, and in receivers it will be either 50 ohms or 300 ohms.

LC MATCHING NETWORKS

LC matching networks are combinations of inductance and capacitance that will allow an input impedance (R1) to be transformed to an output impedance (R2). The common name for these circuits is **antenna tuner** or **antenna tuning unit** (ATU). Several common forms of ATU circuit are shown in Figures 11.3–11.5.

The circuits in Figure 11.3 are for the case where R1 > R2. This situation might occur in vertical antennas, for example, where the feedpoint impedance varies from a few ohms to about 37 ohms, both of which are less than the standard 50 ohms used by the rest of the system. The circuit in Figure 11.3A is one variant of the L-section coupler (two other variants are shown

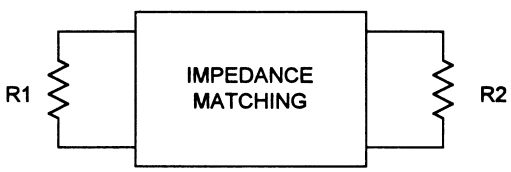


FIGURE 11.2

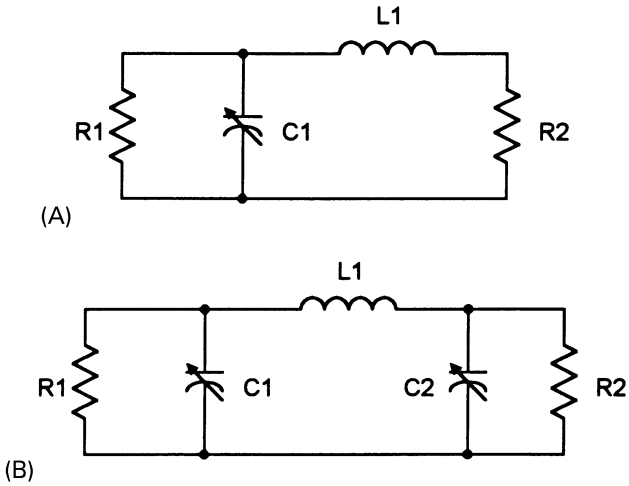


FIGURE 11.3

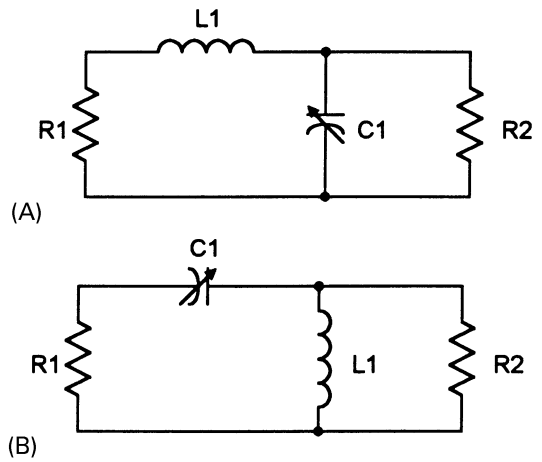


FIGURE 11.4

in Figure 11.4). In this version, the capacitor shunts the line at the input of the network, and the inductor is in series with the line.

Many homebrewed and commercial L-section couplers are designed so that changing jumpers or switch settings allows you to configure the inductor and capacitor in any of the three variants (Figures 11.3A, 11.4A, and 11.4B).

The circuit in Figure 11.4B is the π-network. This type of circuit is often used for impedance transformation when R1 is very much larger than R2 ($R1 \gg R2$). It was once very common as a transmitter output network

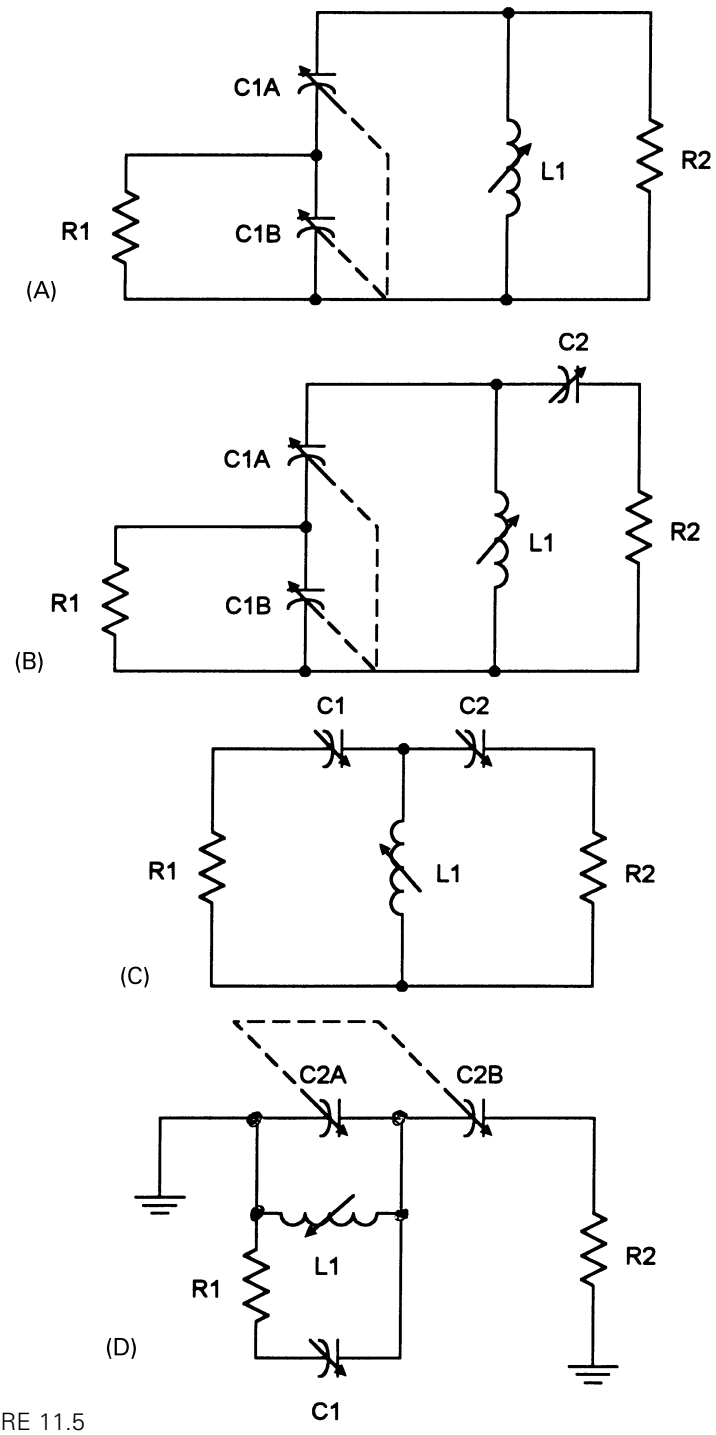

C1A
L1
R2
R1
C1B
(A)
C2
C1A
L1
R2
R1
C1B
(B)
C1
C2
R1
L1
R2
(C)
C2A
C2B
L1
R2
R1
(D)
C1

FIGURE 11.5

because vacuum tube RF power amplifiers needed to match 3000–6000 ohm anode impedances to 50 ohms for the antenna output. The network consists of two shunt capacitors on either side of a series inductor.

A series of four transmatch-style antenna tuning unit circuits is shown in Figure 11.5. These circuits are used in a wide variety of commercially available high-frequency (HF) ATUs. Some of them use dual capacitors, i.e. two-section capacitors in which two separate capacitors share the same shaft. All of the capacitors in these should be wide-spaced 'transmitting' capacitors if the ATU is used as part of a transmitting antenna. For receive-only cases, however, ordinary receiver-style variable capacitors can be used. Although the invidual component values can be calculated, it is common to use 140 or 250 pF for the capacitors, and either 18 μH or 28 μH for the inductor.

The inductor is either a variable or switchable type. Variable inductors are built with a roller mechanism that selects the turns required for any specified value of inductance. However, these inductors can be quite pricey, so many builders opt instead for an air core inductor with several taps at various points along the coil. By selecting the tap one can change the inductance.

The calculation of values for the various style of LC impedance matching network is not terribly difficult if you are trained in mathematics. But if you either prefer the convenience of computer calculations, or are a person who found a first course in algebra and trigonometry a daunting and fearsome experience, then you might wish to take advantage of the software supplement to this book.

QUARTER-WAVELENGTH MATCHING SECTIONS

The quarter-wavelength matching section, also called the Q-section, is made of transmission line (segment 'A' in Figure 11.6). The characteristic impedance of the Q-section line (Z_A) is selected to provide impedance transformation of the load impedance (Z_L) to the characteristic impedance of the line ('B') to the transmitter (Z_B). In some cases, the Q-section is connected directly to the transmitter output (or receiver input), in which case Z_L is transformed to Z_S.

There are two things that you must calculate when making a quarter-wave matching section: the **physical length of the transmission line** and the **characteristic impedance** required to match the load impedance to the source impedance. The physical length to make an electrical quarter wavelength is foreshortened from the physical quarter wavelength due to the velocity factor of the cable. The calculation is

$$L_{\text{meters}} = \frac{75V}{F_{\text{MHz}}} \text{ meters}$$

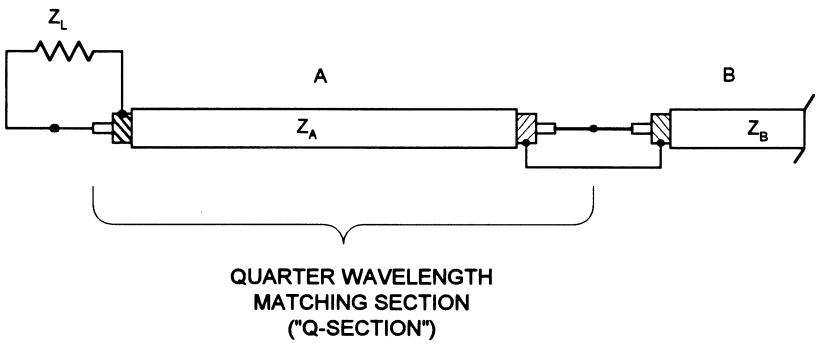


FIGURE 11.6

where L is the length in meters, F_{MHz} is the center operating frequency in megahertz, and V is the velocity factor of the transmission line (either 0.66 or 0.80 for coaxial cable – see Chapter 3).

Calculating the required characteristic impedance requires knowledge of the value of the load impedance and the characteristic impedance of the transmission line to the receiver or transmitter. The calculation is:

$$Z_{\mathrm{A}} = \sqrt{Z_{\mathrm{L}} Z_{\mathrm{B}}}$$

For example, suppose we wish to make a Q-section for matching a 100 ohm impedance (such as found at the feedpoint of a quad loop) to 52 ohm coaxial cable. If the match is not made, then the VSWR will be about 2:1, which for many uses is clearly too high. The calculation is

$$Z_{\mathrm{A}} = \sqrt{(100\,\mathrm{ohms}) \times (52\,\mathrm{ohms})}$$
$$Z_{\mathrm{A}} = \sqrt{5200} = 72\,\mathrm{ohms}$$

The required characteristic impedance is 72 ohms, which is a very good match to 75 ohm coaxial cable. Lest you think this is an example contrived to make the arithmetic work out to a standard impedance found in coaxial cable catalogs (which it is), it is also a very real example that is used for a number of practical antennas. Clearly, though, if the result of this equation is not near enough to some standard coaxial cable to reduce the VSWR to an acceptable level, then some other means of matching the impedance must be found.

The method of Figure 11.6 suffers from the limitations discussed above, namely the possible lack of a readily available characteristic impedance value. However, a much more general case is shown in Figure 11.7. This is the **series matching section** method. Sections L1 and L3 are made of the same type of cable, with the same characteristic impedance. The center section, L2, is made of a cable with a different characteristic impedance,

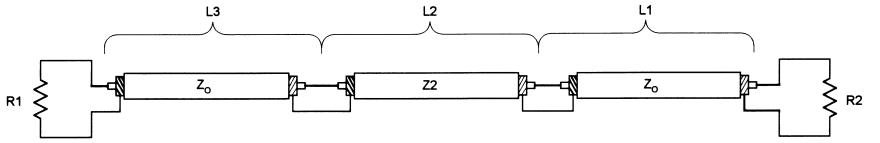


FIGURE 11.7

e.g. 75 ohm coaxial cable or 300 ohm twin-lead. The calculation of the different lengths is a bit daunting, so you are referred to the software supplement to this book in order to make the calculations.

In both the series matching section and the Q-section (which is actually a limited special case of the series section), coaxial cable, twin-lead, or parallel line can be used for the transmission lines, even though only coaxial cable is shown here.

MATCHING STUBS

Another means for effecting the impedance match is to use either a quarter-wavelength or half-wavelength **matching stub** (Figure 11.8). The stub is sometimes mounted at the feedpoint of the antenna, and the transmission line is tapped on the stub at a point (L1) that matches the impedance. In the case shown in Figure 11.8A, a shorting bar is provided at the bottom of the stub. In some cases, the stub is open and the shorting bar is not used.

Another case is shown in Figure 11.8B. In this circuit, the transmission line is connected directly to the antenna, and the stub is used to remove complex impedances by being positioned somewhere between the antenna feedpoint and a point (L1) that is a quarter wavelength down the transmission line from the feedpoint.

The final case is shown in Figure 11.8C. In this scheme, the transmission line is coaxial cable, otherwise this scheme is like Figure 11.8B. The three sections of coaxial cable are connected together using a coaxial 'tee' connector. Typically of the PL-259/SO-239 family, these connectors have three ports that are internally all connected together.

The lengths L1 and L2 depend on the VSWR experienced on the particular antenna, as well as the relationship of Z_L and Z_0. Two cases are seen:

(1) $Z_L > Z_0$:

$$\text{L1} = \arctan\sqrt{\text{VSWR}}\ \text{degrees}$$

$$\text{L2} = \arctan\left(\frac{\sqrt{\text{VSWR}}}{\text{VSWR} - 1}\right)\ \text{degrees}$$

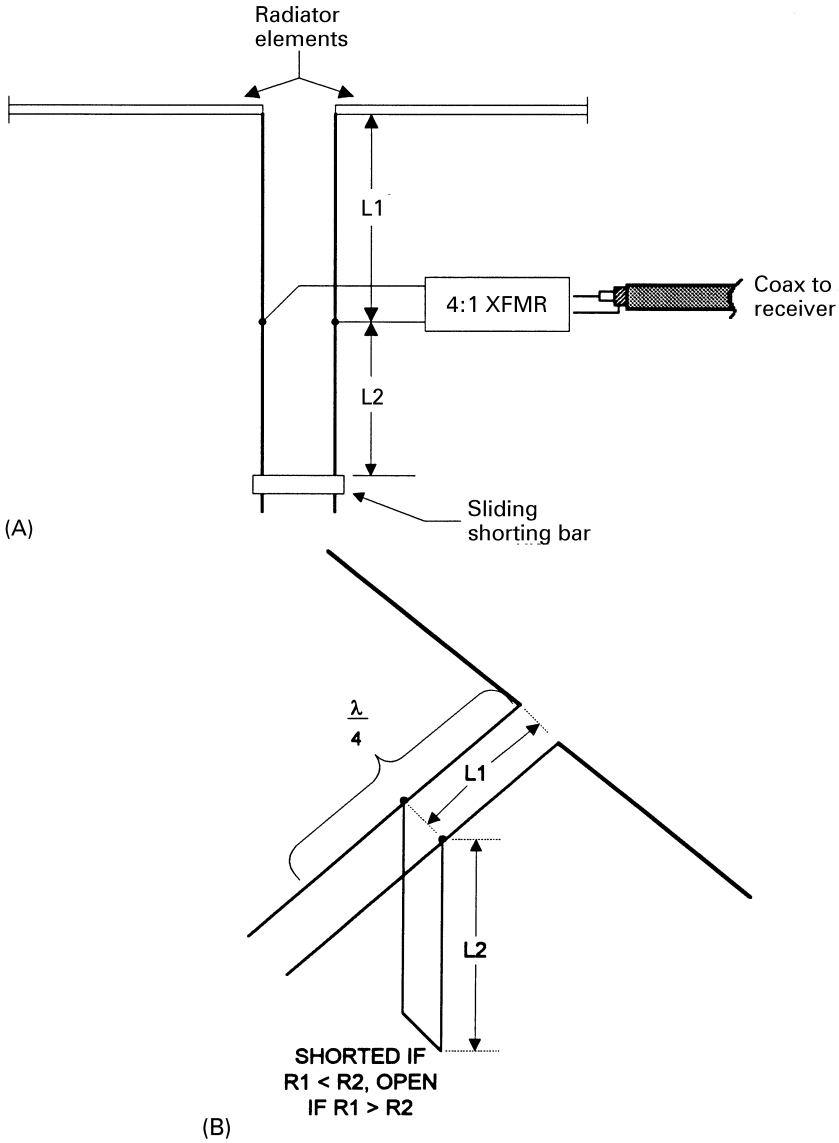
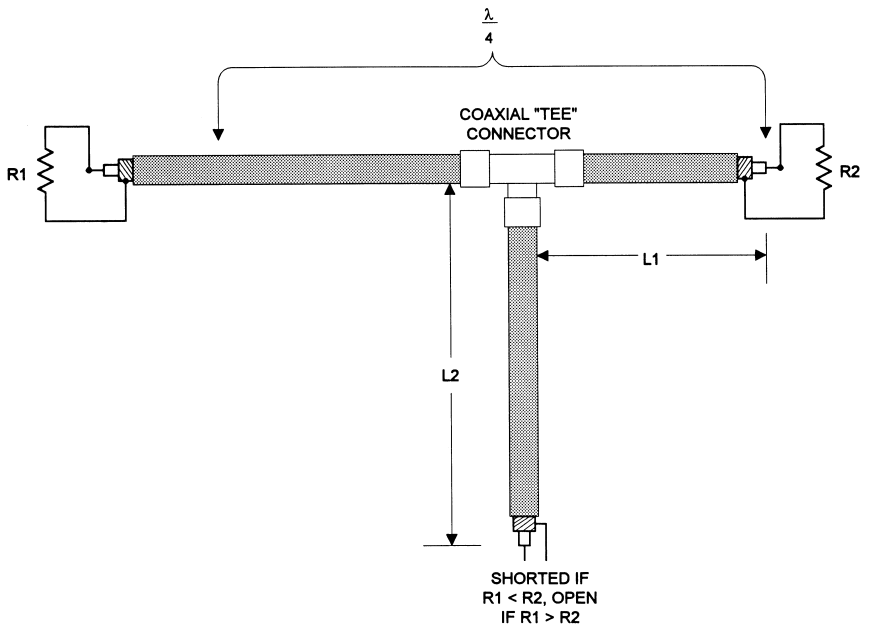


FIGURE 11.8

(2) $Z_L < Z_0$:

$$L1 = \arctan \frac{1}{\sqrt{VSWR}} \text{ degrees}$$

$$L2 = \frac{VSWR - 1}{\sqrt{VSWR}} \text{ degrees}$$



(C)

FIGURE 11.8 (*continued*)

Note that the results of these equations are in degrees. In a transmission line, one wavelength equals 360°, a half wavelength is 180°, and a quarter wavelength is 90°. You can scale the actual length against these values, keeping in mind the velocity factor of the transmission line. Consider an example. Suppose we need to match an antenna that has a 3:1 VSWR on its resonant frequency of 9.75 Mhz using 300 ohm twin-lead ($V = 0.82$). Further, the impedance of the load is 900 ohms (note: VSWR $= 900/300 = 3{:}1$). Because $Z_L > Z_0$, we use the first pair of equations:

$$\begin{aligned} \text{L1} &= \arctan \sqrt{3} \\ \text{L1} &= \arctan\,(1.73) \\ &= 60^\circ \\ \text{L2} &= \arctan\left(\frac{\sqrt{3}}{3-1}\right) \\ \text{L2} &= \arctan \frac{1.73}{2} \\ &= \arctan\,(0.865) \\ &= 40.9^\circ \end{aligned}$$

For a VSWR of 3:1, then we need L1 = 60° and L2 = 40.9°. We can calculate these as a fraction of a quarter wavelength (90°), using

$$\text{Length} = \frac{L \times 75V}{90F_{\text{MHz}}} \text{ meters}$$

where 'Length' is the physical length of the transmission line segment L1 or L2, L is either L1 or L2 expressed in degrees as calculated above, V is the velocity factor of the transmission line, and F_{MHz} is the operating frequency in megahertz.

For the 9.75 MHz cable, using 300 ohm twin-lead ($V = 0.82$), the physical lengths are

$$\text{L1} = \frac{(60)(75)(0.82)}{(90)(9.75)} \text{ meters}$$

$$\text{L1} = \frac{3690}{877.5}$$

$$= 4.2 \text{ meters}$$

$$\text{L2} = \frac{(41)(75)(0.82)}{(90)(9.75)}$$

$$= 2.87 \text{ meters}$$

The matching stub is used on a large number of antennas. Once you get away from the simple half-wavelength dipole fed with 75 ohm coaxial cable, the number of antennas that require some form of impedance matching increases considerably, and the matching stub is used for a lot of different designs.

BALUN AND OTHER BROAD-BAND TRANSFORMERS

There are a number of different transformers used in impedance matching in antenna systems. The term 'BALUN' is used extensively, and it comes from **BALanced-UNbalanced**. Correctly used, then, the term 'BALUN' refers to a transformer that matches a balanced load (e.g. a dipole antenna feedpoint) to an unbalanced load (e.g. coaxial cable). However, it has become common (if erroneous) practice to use 'BALUN' in a generic sense to refer to any of the **broad-band transmission line transformers**. More correctly, some of those are balanced–balanced, so should be called BAL–BAL, and others are of an unbalanced–unbalanced configuration so are called UN–UN. I suppose BALUN sounds more like a word than BAL–BAL or UN–UN (note: some antenna and accessories catalogs do use these terms correctly, but the erroneous usage is also seen).

One of the earliest forms of BALUN transformer is the coaxial cable BALUN shown in Figure 11.9. Both pieces of coaxial cable used in this BALUN transformer are of the same type (e.g. 75 ohm coaxial cable). When

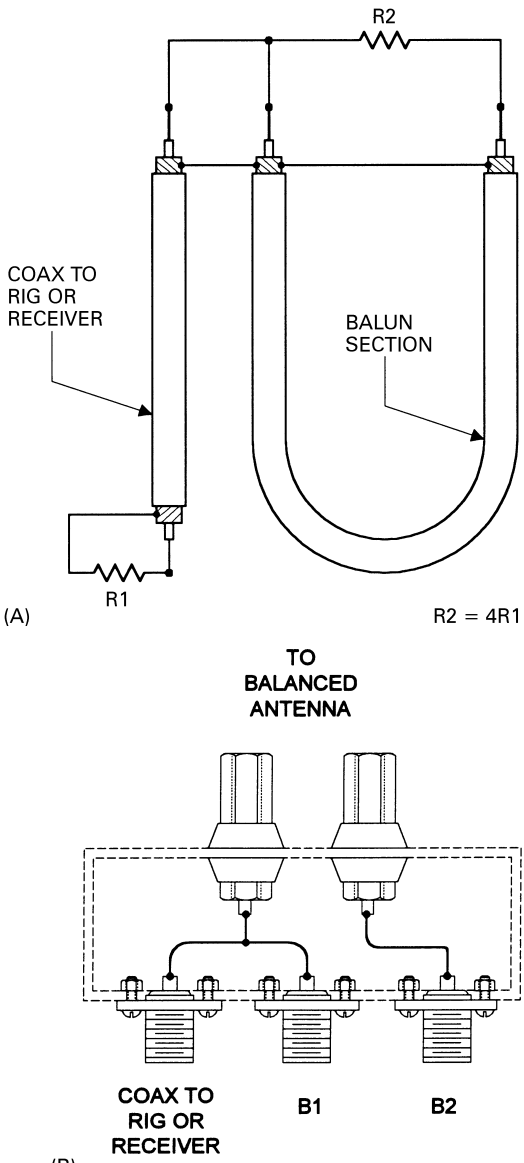


FIGURE 11.9

connected in this configuration, the BALUN transformer produces a 4:1 impedance transformation, which means that a 300 ohm balanced antenna (e.g. folded dipole) will look like a 75 ohm unbalanced load. The coaxial cable to the ham rig or receiver can be of any convenient length. The BALUN section, however, must be a half wavelength long (keeping in mind the velocity factor). The length of the BALUN section is found from:

$$L_{\text{meters}} = \frac{150V}{F_{\text{MHz}}} \text{ meters}$$

where F_{MHz} is the frequency in megahertz and V is the velocity factor (0.66 for polythene coaxial cable, and 0.80 for polyfoam coaxial cable).

A connection box for making the coaxial BALUN is shown in Figure 11.9B. This box is intended for mounting on the antenna center insulator, and should not be used for supporting the antenna unless eye bolts or other more rugged fixtures are provided at the left and right ends. The balanced antenna feedpoint is connected to a pair of five-way binding posts, while the coaxial cable for the run to the rig or receiver and the BALUN sections (B1 and B2) are connected to SO-239 coaxial connectors.

The coaxial BALUN is designed for a specific frequency, but will work over a small margin either side of the design frequency (e.g. typically one HF ham band can be accommodated). But for wide-band operation, you might want to build a broad-band transformer such as those shown in Figure 11.10. Note that some of these transformers only show one core symbol (e.g. Figure 11.10A). Those transformers have all the windings on the same core. The dots show the phase sense of the windings and indicate the same end of the winding).

The two most common forms of BALUN transformer are those in Figures 11.10A and 11.10B. The version in Figure 11.10A has no impedance transformation, and is usually referred to as a 1:1 BALUN. The transformer in Figure 11.10B, on the other hand, offers a 4:1 impedance transformation, so is equivalent to the coaxial BALUN shown in Figure 11.9A. The transformers in Figures 11.10C and 11.10D are both UN–UNs. The configuration in Figure 11.10 produces a 9:1 impedance transformation, while that in Figure 11.10D produces a 16:1 transformation.

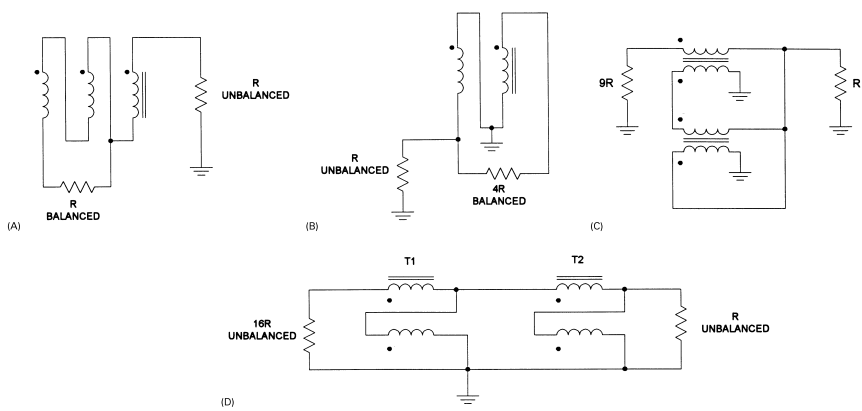


FIGURE 11.10

The construction of coil BALUNs and broad-band transformers is shown in Figure 11.11. The transformer shown in Figure 11.11A is wound on a toroid core made of either powdered iron or ferrite material. The toroid is doughnut shaped. It has the interesting attribute of containing the magnetic field to its own geometry, so has little interaction with its environment. This fact means that it will work like the book says more often than certain other transformer core configurations.

Note how the wires are wound on the toroid core. They are kept paired and lay next to each other. In this manner they are wound together as if they were only one piece. When two wires are used, this form of winding is called **bifilar winding**. When three wires are wound together in this manner, a **trifilar**

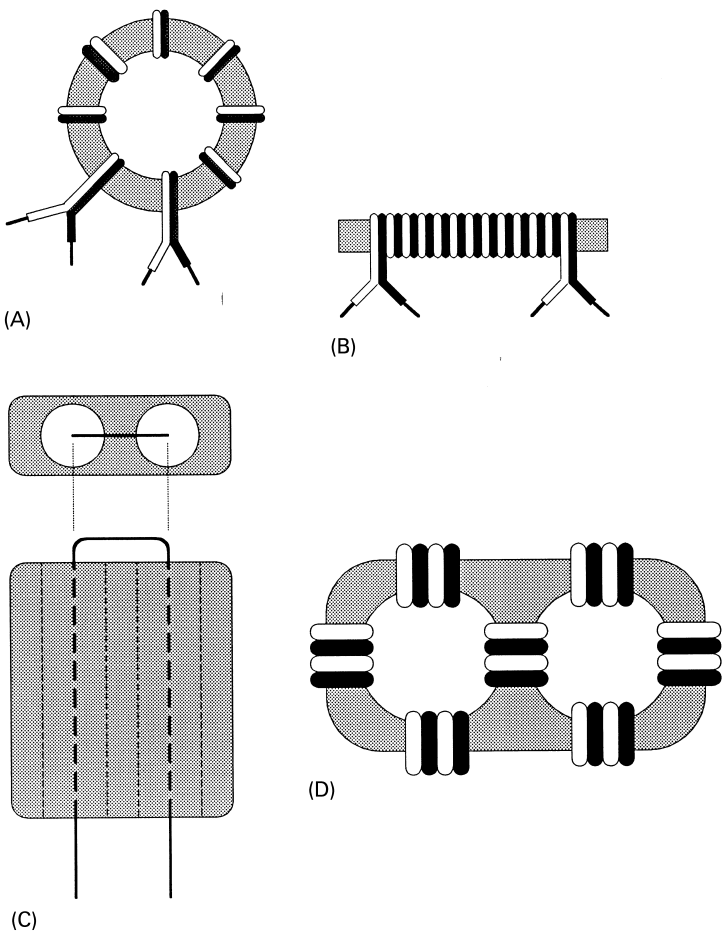


FIGURE 11.11

winding is produced. The bifilar method is used to wind the transformer in Figure 11.10B, while trifilar winding is used for that in Figure 11.10A.

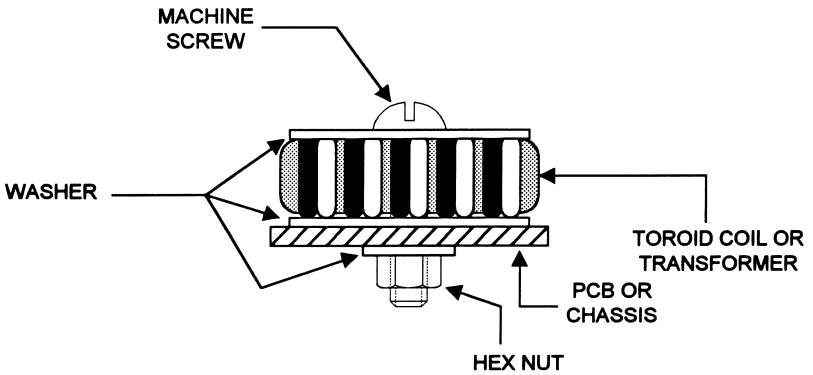
The solenoid winding method is shown in Figure 11.11B. The core can be either air (in which case a coil form is needed), or a ferrite rod (as shown). Again we see the use of either bifilar or trifilar winding, depending on the nature of the transformer being made.

The so-called bazooka BALUN core is shown in Figures 11.11C and 11.11D, using two different winding styles. In the style of Figure 11.11C, the wire is passed through both holes to form a loop ('internal winding'). Counting the number of turns is a little different than one might suppose. The case shown in Figure 11.11C is one turn, even though many people erroneously assume that it is half a turn. If one end of the wire is passed through both holes one more time, then there are two turns present. Both the primary and secondary windings can be wound in this same manner, laying one over the top of the other (the primary is usually laid down first). The case shown in Figure 11.11D shows an end view of the bazooka BALUN core. In this case, several turns are wound in both the internal and external winding styles. These two styles can be intermixed on the same form, but wherever possible you are advised to use the internal winding mode preferentially.

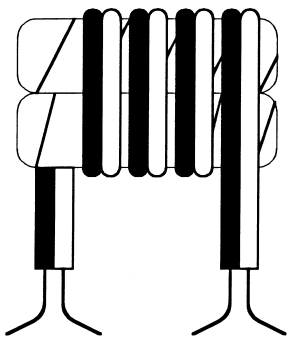
Figure 11.12A shows how a toroid core inductor or transformer is mounted on either a printed circuit board (PCB) or metal chassis. Fiber or nylon washers are used to secure and protect the toroidal core, and nylon or other non-metallic fasteners (machine screw and hex nut) are used to keep it in place. It is important to use non-metallic fasteners to keep from interfering with the operation of the transformer. Only in the case of the largest toroids (>5 or 6 cm in diameter) are metal fasteners usable, and even then there is some bad effect.

The scheme in Figure 11.12B is used for transmitter ATUs and similar applications where the power is higher. Two or more 5 cm or larger toroids are stacked one on top of the other. Each toroid core is first wound with fiberglass tape to insulate it from the other core. After the cores are wrapped with tape and stacked on top of each other, a final layer of tape can be added to keep the whole assembly stable. The bifilar or trifilar windings are then laid down on the stacked cores.

A number of manufacturers offer BALUNs, BAL–BALs, and UN–UNs, in both voltage and current configurations. Some are designed to replace the center insulator of an antenna such as the dipole. Others are intended for mounting elsewhere. The transformer in Figure 11.13 is intended for use on vertical antennas. The feedpoint impedance of the quarter-wavelength vertical will vary from something less than 5 ohms to about 37 ohms, depending on the installation. The transformer of Figure 11.13 is an UN–UN that has a single attachment point for 52 ohm coaxial cable to the rig or receiver.



(A)



(B)

FIGURE 11.12

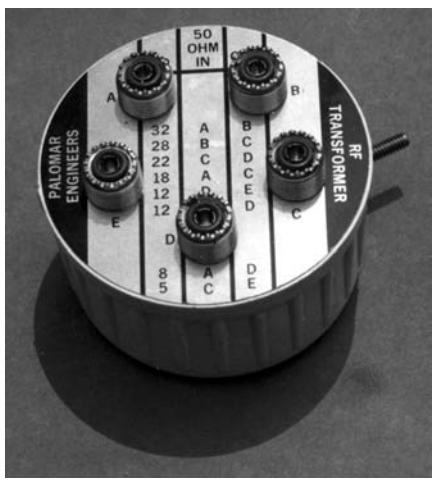


FIGURE 11.13

The other four coaxial connectors are various load impedances that will be transformed to 52 ohms if that port is used.

COMMERCIAL ANTENNA TUNERS

An 'antenna tuner' located at the transmitter end of the transmission line will not effect a good impedance match to the antenna, but it can mitigate the effects of a high VSWR on the transmitter. Even a single-band resonant antenna will exhibit a high enough VSWR at the ends of the band to upset the operation of the transmitter that is equipped with automatic load control (that is, all modern solid-state rigs). Although antenna tuner construction is well within the abilities of most do-it-yourself 'homebrewers,' it is also likely that most will opt instead for a ready-built ATU. A number of reasonably priced models are available on the market.

Figure 11.14 shows the MFJ-956 receiver S/M/L wave preselector/tuner. This device places a series resonant circuit in series with the antenna downlead before it goes to the receiver. A single variable capacitor (365 pF maximum) is in series with an inductance. Four different switch-selected inductance values are provided (2.5 mH, 330 μH, 22 μH, and 1 μH) to cover the range 150 kHz to 30 MHz. When the inductor switch is in the BYP ('bypass') position, the preselector/tuner is effectively out of the circuit, and the antenna–receiver combination works as normal. In the GND ('ground') position, the antenna input of the receiver (but not the antenna) is grounded.

Tuners for transmitting use are shown in Figures 11.15 and 11.16. The tuner in Figure 11.15 is the MFJ-949E Deluxe Versa Tuner II, while that in Figure 11.16 is the MFJ-986 Differential Tee tuner. These tuners are a bit larger than the standard receiver-only tuner, but are nonetheless just as

FIGURE 11.14

FIGURE 11.15

FIGURE 11.16

useful to the receiver owner as well. The only thing that will not work for receiver owners is the VSWR metering system. These are based on the same technology as RF wattmeters, so require the output of a radio transmitter for excitation. Both models allow more than one antenna (although only one at a time!) to be switch selected.

CHAPTER 12

Simple antenna instrumentation and measurements

It is very difficult to get an antenna working properly without making some simple measurements. Although it would be nice to make pattern measurements, those are beyond the reach of almost all of us. On the other hand, the 'I worked [or heard] a guy on the other side of the world' type of measurement tells us little or nothing. During the peak of the sunspot cycle a breath of hot air on the antenna can be picked up on all continents.

There are some things that can be measured, however. For example, the voltage standing wave ratio (VSWR) and the resonant frequency of the antenna are readily accessible. It is also possible to measure the impedance of the antenna feedpoint. Hams can measure the VSWR either with a special VSWR meter (often built into transmitters or antenna tuning units), or by using a radio frequency (RF) wattmeter.

By stepping through the band and testing the VSWR at various frequencies, one can draw a VSWR curve (Figure 12.1) that shows how the antenna performs across the band. The resonant frequency is the point where the VSWR dips to a minimum. You can use the resonant frequency to figure out whether the antenna is too long (resonant frequency lower than the hoped-for design frequency), or too short (resonant frequency above the design frequency).

But resonant frequency and VSWR curves are not the entire story because they do not tell us anything about the impedance presented by the antenna. One cannot get the VSWR to be 1:1 unless the antenna impedance and transmission line impedance are the same. For example, a dipole has a nominal textbook impedance of 73 ohms, so makes a very good match

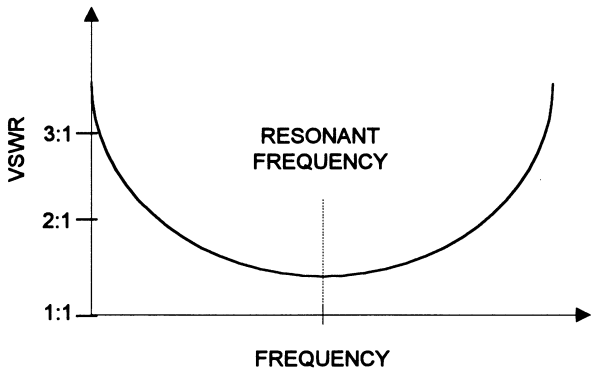


FIGURE 12.1

to 75 ohm coaxial cable. But the actual impedance of a real dipole may vary from a few ohms to more than 100 ohms. For example, if your antenna exhibits a feedpoint impedance of (say) 25 ohms, using 75 ohm coaxial cable to feed it produces a VSWR = 75/25 = 3:1. Not too great. Measuring the feedpoint impedance is therefore quite important to making the antenna work properly.

There are any number of instruments on the market that will aid in making antenna measurements. Some of them are quite reasonably priced (or can be built), while others are beyond the reach of all but the most ardent and well-endowed enthusiasts. In this chapter we will look at RF power meters, VSWR meters, dip meters, noise bridges, impedance bridges, and a newer breed of more universal instrument called the SWR analyzer. We will also look at a few accessories that are useful in the measurement or adjustment of radio antennas (e.g. coaxial switches, attenuators, and dummy loads).

RF POWER METERS AND VSWR METERS

The RF power meter is designed to indicate the actual value of RF power (in watts). VSWR meters are similar, but are either calibrated in VSWR units or have a selectable scale that is so calibrated. Some RF wattmeters measure the voltage on the transmission line, while others measure the current flowing in the line. If things are done right, then these approaches work nicely. But sometimes those methods can be fooled. A better system is to measure both the current and the voltage and to display their product. Although this task sounds like a formidible instrumentation job, it is actually the way one of the most popular RF power meters works. The Bird Electronics Model 43 (Figure 12.2) is one of the oldest and most popular professional-grade instruments. It uses a specially designed internal transmission line segment

FIGURE 12.2

with sampling loops inserted in the form of plug-in 'elements' (the round object with an arrow in Figure 12.2). The elements are customized for the RF power level and frequency range. The arrow on the element indicates the direction of the power flow. When the arrow points to the antenna feedline, the instrument is measuring the forward RF power, and if pointed toward the transmitter it is measuring the reflected RF power.

Although the Model 43 type of meter will provide the RF power measurements more accurately under a wider range of conditions, a ham radio style VSWR meter/RF power meter will work well (and measure accurately) under most ham conditions. In addition, these instruments will provide a measure of the VSWR (with an RF wattmeter you need to calculate the VSWR). Figure 12.3 shows two typical ham-style VSWR/RF power meters. The instrument in Figure 12.3A is small and simple, but quite effective. It will measure forward and reflected power over the entire high-frequency (HF) band, as well as the VSWR. To measure the VSWR, set the full-scale/SWR switch (marked 'FS' and 'SWR') to the FS position, key the transmitter, and adjust the knob to achieve a full-scale deflection of the meter pointer. When the FS/SWR switch is set to the SWR position, it will read the VSWR. These instructions, by the way, fit a large number of different instruments, by many manufacturers, in this same class.

The instrument in Figure 12.3B is for the very high/ultrahigh frequency (VHF/UHF) bands. It does not have to be set as does the meter above. The

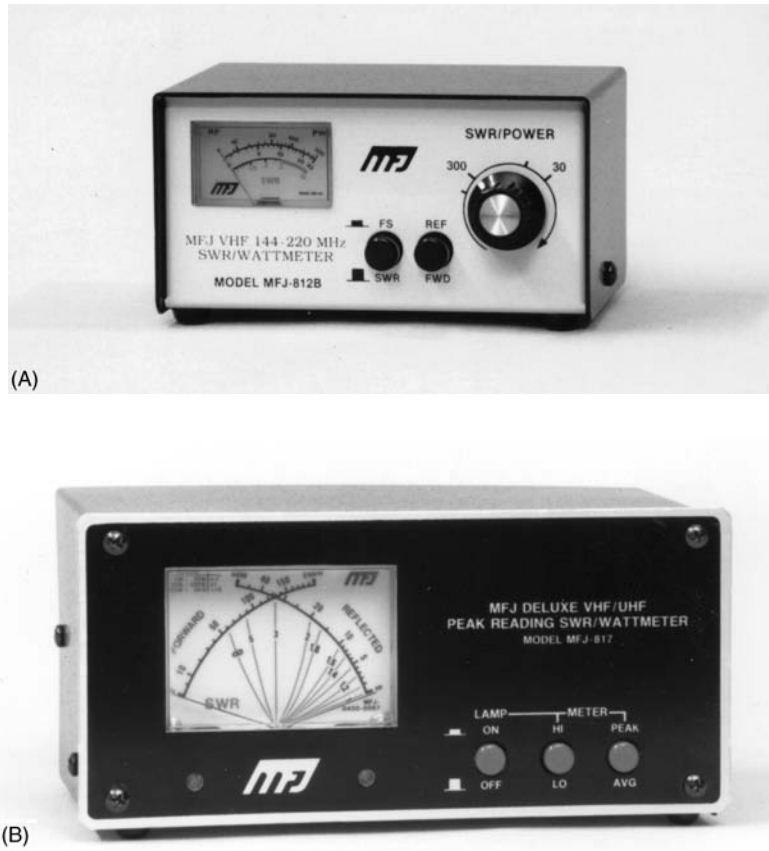


FIGURE 12.3

reason is that it uses the 'crossed-needles' method of displaying forward and reverse power. Lines of intersection with the two scales form a species of nomograph that reads out the VSWR from observation of the point where the needles cross. This instrument also includes a **peak reading** function. Most passive RF power meters inherently measure the average RF power because of the inertia of the meter movement. But with active circuitry, i.e. circuits that require direct-current power to operate, it is possible to sample and hold the peaks of the RF waveform. This meter then becomes more useful for measuring continuous-wave and amplitude-modulated waveforms (including SSB). Some authorities claim that the peak reading meter is preferred for SSB operation – and I concur.

DIP METERS

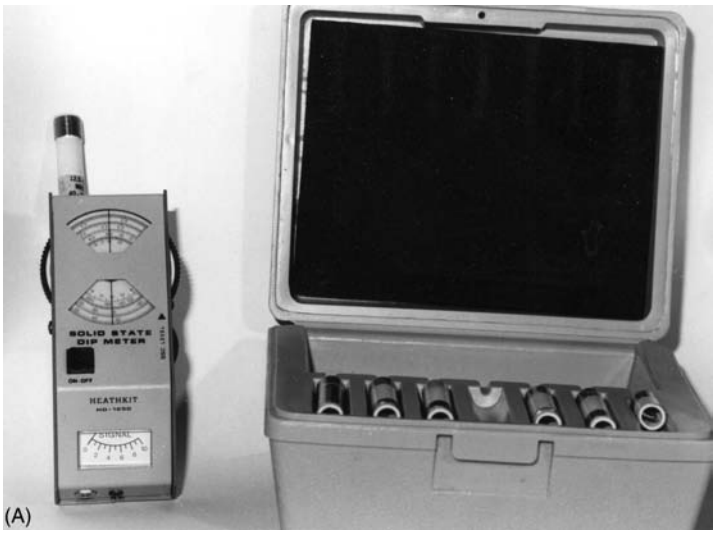
The **dip meter** works because a resonant circuit will draw power from any nearby field oscillating at its resonant frequency. An antenna that is cut to a specific frequency is resonant, so acts like an LC resonant circuit. At one time, these instruments were called **grid dip meters** because they were vacuum tube ('valve') operated. The metering monitored the grid current to detect the sudden decrease when the instrument was coupled to an external resonant circuit. Today, some people still use the archaic term 'grid dipper' even though dip meters have not had grids for nearly 30 years.

Figure 12.4A shows a typical dip meter. It consists of an LC oscillator in which the inductor is protruding outside the box so that it can be coupled to the circuit under test. A large dial is calibrated for frequency, and a meter monitors the signal level. Coils for additional bands are stored in the carrying case.

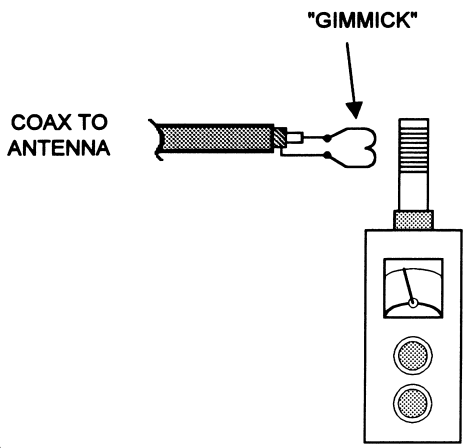
The frequency dial of the dip meter bears some additional comment. It typically covers a very wide range of frequencies, yet the dip we need is very sharp. The calibration is none too accurate. As a result, it is wise to find the dip, and then use a receiver to measure the operating frequency while the dip meter coil is still coupled to the circuit being tested. This operation is a neat trick unless you are possessed with good manual dexterity. It is necessary to measure the frequency while the instrument is coupled because this type of LC oscillator changes frequency markedly when the loading of the coil changes.

The proper method for coupling the dip meter to an antenna transmission line is shown in Figure 12.4B. Construct a 'gimmick', i.e. a small coil of one to three turns, of sufficient diameter to slip over the coil protruding from the dip meter. Either slip the dipper coil inside the gimmick, or bring it into close proximity. Energy is transferred from the dipper to the gimmick and thus to the antenna. When the dipper is tuned to the resonant frequency of the antenna, a sudden drain is put on the dipper, and the signal level drops sharply. The frequency of this dip tells you the resonant frequency of the antenna.

The dipper must be tuned very slowly – very, very slowly – or you will miss the dipping action of the meter. Tuning too fast is the principal reason why many newcomers fail to make the dipper work. To make matters more confusing, the meter needle will drift all over the scale as the instrument is tuned. This occurs because it is natural for LC-tuned oscillators to produce different signal level at different frequencies (generally, but not always, decreasing as frequency increases). The effect to watch for is a very sharp drop in the meter reading.



(A)



(B)

FIGURE 12.4

ANTENNA IMPEDANCE (OR 'RESISTANCE') BRIDGES

While it is necessary to measure the VSWR of an antenna to see how well it is working, the VSWR alone is insufficient to optimize performance. If you do not know the feedpoint impedance at resonance (minimum VSWR), then you cannot do much to correct a problem. Several instruments are available that will measure the feedpoint resistance of the antenna (it is called 'impedance' in some user manuals, but it is really only the resistive component of

impedance). The instruments in Figure 12.5 are examples of antenna resistance meters. The instrument in Figure 12.5A is based on the Wheatstone bridge circuit. It uses a variable resistor (connected to the thumbwheel resistance dial) as one leg of the bridge, and the antenna

(A)

(B)

FIGURE 12.5

resistance another leg (fixed resistors form the remaining two legs of the classic Wheatstone bridge). This model is passive, and requires an external signal source. There is an internal amplifier that allows you to use a signal generator, or the amplifier can be bypassed if you wish to drive it with a transmitter signal.

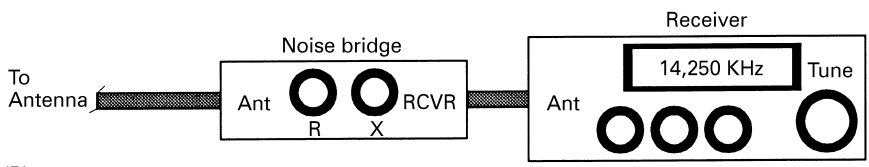
The instrument in Figure 12.5B is a little different. It is basically a sensitive VSWR meter with a built-in signal generator. Knowing the system impedance it can thereby calculate the resistance from knowledge of the VSWR.

RF NOISE BRIDGES

The **noise bridge** (Figure 12.6A) is an interesting antenna instrument that is used in conjunction with a radio receiver to measure the antenna feedpoint resistance, approximate the reactance component of the impedance, and find the resonant frequency. These instruments use white noise ('hiss' is what it sounds like on the receiver speaker) to generate a wide spectrum of radio frequencies. When the receiver is used to monitor the noise, a null is noted at the resonant frequency.

Figure 12.6B shows how the noise bridge is connected between the antenna transmission line and the antenna input connector of the receiver. The line between the bridge and the receiver can be any length, but it is wise

(A)



(B)

FIGURE 12.6

to keep it short. One reader of my columns wrote in to take exception to the 'short as possible' requirement because it is not strictly necessary in a *perfect world*. My own experience says differently, especially at higher frequencies where the line is a significant fraction of a wavelength. A meter or so should not affect any HF measurement, however.

The line to the antenna should either be as short as possible (preferably zero length), but in the real world, where one has to be more practical than theoretically pure, it is sufficient to make the transmission line an integer multiple of an electrical half wavelength long (the physical length is shorter than the free space half wavelength because of the velocity factor). The impedance hanging on the far end of the half wavelength $\times N$ line is repeated every half wavelength. A consequence of this effect is that the impedance measured at the transmitter or receiver end is the same as at the antenna end.

To use the noise bridge, set the resistance (R) and reactance (X) controls to mid-scale (X = 0). Adjust the receiver frequency to the expected resonant frequency. At this point you can do either of two different measurement schemes.

Method 1

First, you can set the R-control of the noise bridge to the desired antenna impedance, and then tune the receiver until a dip in the noise occurs (as heard in the output or noted on the S-meter). For example, suppose you have a dipole cut for 11.75 MHz, and installed such that the expected feedpoint impedance should be around 70 ohms. It is fed with a half wavelength of polyfoam dielectric 75 ohm coaxial cable ($V = 0.80$). The cable length is $(150)(0.8)/11.75 = 10.2$ m. The noise bridge X-control is set to 0, and the resistance control is set to the letter that represents 70 ohms (given in the calibration manual). This resistance setting can also be determined experimentally by connecting a 70 ohm resistor across the antenna terminal and adjusting the R-control for the minimum noise. When the noise bridge is set, then adjust the receiver for minimum noise. The frequency at which the dip occurs is the actual resonant frequency. I used this method to investigate a vertical cut for 21.25 MHz, only to find the actual resonance was 19.2 MHz, which explained the high VSWR.

Method 2

The other method is to set the receiver to the design resonant frequency of the antenna. You then adjust the R-control for a dip in the noise level. The value of the resistance is the feedpoint impedance of the antenna. You can either match that impedance, or adjust the antenna to bring the actual resonant frequency closer to the design frequency.

VSWR ANALYZERS

The basic premise in this section is that the instruments used must be accessible to people who do not have a ham radio or professional radio operator's license. Some of the instruments discussed above meet that requirement, but a relatively new breed of instrument called the SWR analyzer provides a lot of capability to the short-wave listener, scanner operator, and ham radio operator alike. It uses a low-power RF signal generator and some clever circuitry to measure the VSWR of the antenna. One model also measures the feedpoint resistance.

The simple version shown in Figure 12.7 is for the low VHF band (up to and including the 6 m ham band). It is a hand-held, battery-powered instrument. The meter reads the VSWR of the antenna at the frequency set by the TUNE dial.

A somewhat more sophisticated instrument, the MFJ Enterprises Model MFJ-259, is shown in Figure 12.8. The front panel has two meters, SWR and RESISTANCE. The SWR meter is calibrated up to 3:1, with a little uncalibrated scale to indicate higher SWRs. The RESISTANCE meter is calibrated from 0 to 500 ohms, which is consistent with the SWR range. Two controls on the front panel are TUNE and FREQUENCY (MHz) (a

FIGURE 12.7

FIGURE 12.8

bandswitch). The MFJ-259 has a digital frequency meter to measure the operating frequency of the internal oscillator. This frequency counter can also be used to measure the frequency of external signal sources (DO NOT connect the counter to the output of a transmitter – the instrument will be destroyed). The top end of the MFJ-259 has a number of controls and connectors. An SO-239 'UHF' style coaxial connector is provided for the antenna connection. A BNC coaxial connector is provided to apply an external signal to the frequency counter, while a pushbutton INPUT switch is available to switch the counter from internal to external signal sources. Another pushbutton switch is used to set the gate timing of the counter (a red LED on the front panel blinks every time the gate is triggered). The tuning is from 1.8 to 174 MHz, while the counter will measure up to 200 MHz.

The MFJ-259 will work from an external 12 V DC source, or from an internal battery pack consisting of eight size-AA standard cells. MFJ recommends that either alkaline or rechargeable batteries, rather than ordinary zinc–carbon cells, be used in order to reduce the possibility of leakage that could damage the instrument (this is good practice in all battery-powered instruments). I have a homebrewed battery pack that uses eight size-D nickel cadmium batteries (4 A-h rating) that can be recharged from a 12 V DC power supply, and it works well with the MFJ-259.

Unlike many lesser SWR meters, this instrument is not fooled by antennas that have impedances consisting of both resistance and reactance elements. An example in the manual demonstrates an impedance of 25 + j25 ohms (i.e. R is 25 ohms and reactance, X, is also 25 ohms). When connected to a 50 ohm load one might be tempted to think the VSWR is 1:1, and some cheaper meters will so indicate. But the actual SWR is 2.6:1, which is what the MFJ-259 will read.

The resistance measurement assumes a resistive load (i.e. the measurement is made at the resonant frequency of the antenna), and is referenced to 50 ohms. The VSWR and resistance measurements should be consistent with each other. If the VSWR is 2:1, then the resistance should be either 100 ohms (100/50 = 2:1) or 25 ohms (50/25 = 2:1). If the resistance is not consistent with the VSWR reading, then you should assume that the impedance has a significant reactive component, and take steps to tune it out.

In addition to antenna measurements, the MFJ-259 is equipped to measure a wide variety of other things as well. It will measure the velocity factor of transmission line, help in tuning or adjusting matching stubs or matching networks, measure capacitance or inductance, and measure the resonant frequency of LC networks.

Figure 12.9 shows an MFJ-249 meter (similar to the MFJ-259, but without the resistance measurement) equipped with the MFJ-66 dip meter adapter. It can be used to make the MFJ-249 or MFJ-259 work in the same manner as a dip meter. Using this adapter allows you to measure the resonant frequency of tank circuits using the dipper approach, as well as to measure things such as the coefficient of coupling between two LC circuits, transformers, and other radio circuits.

DUMMY LOADS (ARTIFICIAL AERIAL)

The **dummy load**, or **artificial aerial** as it is also called, is a resistor used to simulate an antenna when adjusting a transmitter or other radio apparatus. These devices consist of a non-inductive, 50 ohm resistor inside a shielded enclosure.

The dummy load offers several benefits over regular antennas. First, it is a constant resistance over the entire frequency range, and, second, it does not present any appreciable reactance. Third, and perhaps most important, the dummy load provides the ability to conduct tests off the air where you will not cause television interference, broadcast interference, or interference to other users of the test frequency that you select. Besides, it is illegal and just plain rude to radiate when you do not have to!

One use for dummy loads is to adjust antenna tuners. You can connect the dummy load to the antenna output of the tuner, and then adjust the

FIGURE 12.9

controls for best VSWR at a number of frequencies. By recording the knob settings, you can 'rough in' the antenna tuner when changing frequency.

Several types of dummy load are shown in Figures 12.10 and 12.11. The version in Figure 12.10 is a small 5 W HF model intended for measuring the output of CB transmitters. I have used this particular dummy load for adjusting a wide range of RF projects and instruments over the years. To make accurate power measurements, the dummy load is placed at the ANT connector of the RF power meter, and the transmitter connected to the other end. The power reading is not obscured by transmission line effects or the reactance that a real antenna might present.

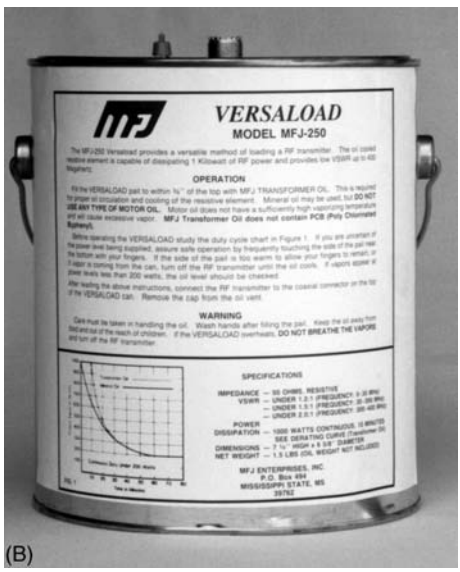
The dummy loads shown in Figure 12.11 are intended for ham radio power levels. The model in Figure 12.11A is air cooled, and operates at RF power levels up to 1500 W, over a range of 1–650 MHz. The version in Figure 12.11B is oil cooled, and will handle up to 1000 W of RF power.

COAXIAL SWITCHES

A coaxial switch (Figure 12.12) is used to allow a receiver or ham radio set to use any of several antennas (models with up to 16 ports are available, but this one is a two-port model). The common connector is for the receiver or transmitter, while the two antennas are connector to the 'A' and 'B' ports,

FIGURE 12.10

(A)



(B)

FIGURE 12.11

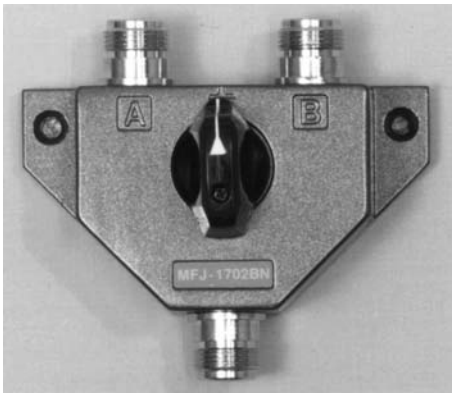


FIGURE 12.12

respectively. Alternatively, one can turn the switch around backwards (it is bidirectional, after all), and use the same antenna on two different receivers or transmitters.

The use of the coaxial switch in antenna measurement is in comparing the antenna being tested with either another antenna or a dummy load. The kind of off-the-air checks that amateurs and short-wave listeners can make are notoriously inaccurate, but can be made a lot more useful by making comparisons with known antennas. For example, a friend of mine, the late Johnnie H. Thorne (K4NFU/5) had an antenna farm in Texas (and it did seem that he grew antennas, judging from the number he had). He kept a standard dipole, optimally installed and cut for 20 m, and made all of his test designs for the same frequency. He would compare new designs to the dipole by switching back and forth while monitoring the signal strength on the receiver S-meter. He could also compare two different antennas by comparing them against each other or against the dipole.

STEP ATTENUATORS

The **step attenuator** (Figure 12.13) is a precision instrument that provides highly accurate levels of attenuation in steps of 1 dB, 2 dB, 3 dB, 5 dB, 10 dBV, and one or more 20 dB settings. You can use these to calibrate instruments, or measure signal levels. One use is to make comparison measurements of two signal sources, two antennas, etc.

For example, suppose you want to measure the gain of a new antenna relative to a dipole (assuming that both are cut for the same frequency). You would pick or provide a distant signal on the resonant frequency of the antenna. The attenuator is inserted into the feedline path of the test antenna (Figure 12.14). The signal strength is then measured with the dipole in the

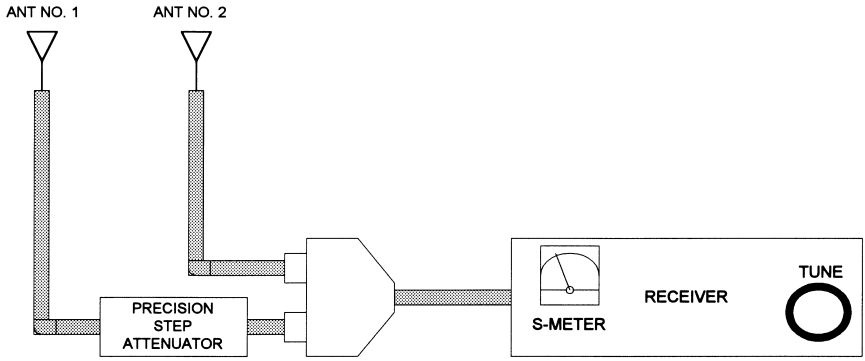


FIGURE 12.13

FIGURE 12.14

circuit (see the coaxial switch discussion above), and the reading of the S-meter noted. If you have an RF sensitivity control, then adjust it to a convenient point on the meter dial, like S-9 or some other marked spot. Next, cut to the new antenna and note the signal strength. If it increased, then select the various levels of attenuation until the new signal level is the same as the initial signal level. The set level of attenuation is equivalent to the gain difference between the antennas. If the new signal level is lower than the initial signal level, then the new antenna has less gain than the old antenna. Put the attenuator in line with the comparison antenna, rather than the test antenna.

CHAPTER 13

Getting a 'good ground'

Because this book is about radio antennas it is necessary to also spend some time on the topic of antenna grounds. The effectiveness and safety of a radio antenna often hangs on whether or not it has a good radio frequency (RF) ground. Poor grounds cause most forms of antenna to operate at less than best efficiency (sometimes a whole lot less). In fact, when transmitting it is possible to burn up between 50 and 90% of your RF power heating the ground losses under the antenna, instead of propagating into the air. Ground resistances can vary from very low values of 5 ohms, up to more than 100 ohms (5–30 ohms is a frequently quoted range). RF power is dissipated in the ground resistance, so is not available for radiation. The factors that affect the ground resistance include the **conductivity of the ground**, its chemical **composition**, and its **water content**.

The ideal ground depth is rarely right on the surface, and, depending on the local water table level, might be a couple of meters or so below the surface. It is common practice among some amateur radio operators to use the building electrical ground wiring for the RF antenna ground of their station. Neglecting to install an outdoor ground that will properly do the job, they opt instead for a single connection to the grounded 'third wire' in a nearby electrical outlet. Besides being dangerous to work with, unless you know what you are doing, it is also a very poor RF ground. It is too long for even the low RF bands, and radiates RF around the house in large quantity. Stations that use the household electrical wiring as the radio ground tend to cause television interference, and other electromagnetic interference both in their own building and in nearby buildings.

Other people use the plumbing piping in the house, but this is also a problem. While the cold water pipe entering older houses is metal, and is buried for quite a distance underground, it is also a single conductor, and is often made of a metal that is not optimal for radio work. Also, in modern buildings, the main water service and the distribution pipes in the plumbing system of the building may well not be metal at all, but rather PVC or some other insulating synthetic material. Even if you get a metal water pipe, if it is the hot water pipe then no effective ground exists. There is also a very severe danger of hooking your ground to the gas service line, which could *cause a fire or explosion*. In the USA, gas pipes are usually painted black, or are unpainted, but enough water pipes also fit this description that it is good practice to avoid the pipes in a building when searching for a radio grounding solution.

TRADITIONAL RADIO GROUNDING SOLUTIONS

The goal in making a radio ground is to provide a low resistance path to the earth for radio signals. An additional goal is to make a path to the earth for lightning that strikes the antenna, or even nearby (see Chapter 2 for the use of lightning arrestors in antenna lines).

We can reduce the ground resistance by either altering the composition of the earth surrounding the ground point, or by using a large surface area conductor as the ground point. Figure 13.1 shows the traditional ground rod used by small radio stations, including amateur and receive-only stations. Use either a copper (or copper-clad steel) rod at least 2–3 m long. Electrical supply houses, as well as amateur radio and communications equipment suppliers, also sell these ground rods. Do not use the non-clad steel types sold by some electrical supply houses. They are usable by electricians when making a service entrance ground in your home or workplace, but RF applications require the low skin resistance of the copper-clad variety. The rod need not be all copper, because of the skin effect forcing the RF current to flow only on the outer surface of the rod. Try to use a 2.5 m or longer rod if at all possible, because it will work better than the shorter kind. Do not bother with the small television antenna 100–150 cm ground rods; they are next to useless for high-frequency (HF) radio stations. Drive the ground rod into the earth until only a few centimeters remain above the surface. Connect a heavy gage ground wire from your station to the ground rod (some people use heavy braid or the sort once used for the return connection from the battery to the body of an automobile; an additional source is salvaged braid from the heavier gages of coaxial cable). The ground wire should be as short as possible. Furthermore, it should be a low-inductance conductor. Use either heavy braid (or the outer conductor stripped from RG-8 or RG-11 coaxial cable), or sheet copper. You can buy

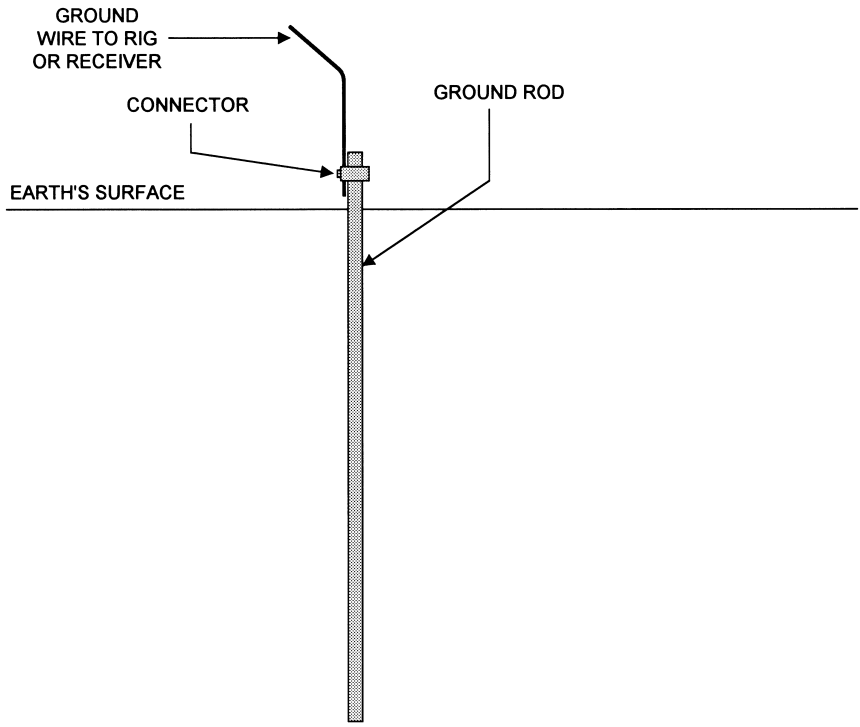


FIGURE 13.1

rolls of sheet copper from metal distributors in widths from 10 cm up to about 50 cm. Some amateurs prefer to use 17.5 cm-wide roofing foil that is rated at a weight of 1.5 kg/m. Sweat solder the ground wire to the rod. You can get away with using mechanical connections like the electricians use, but will eventually have to service the installation when corrosion takes its toll. I prefer to use soldered connections, and then cover the joint with either petroleum jelly or acrylic spray lacquer.

Another alternative is to use a copper plumbing pipe as the ground rod. The pipe can be purchased in 2–5 m lengths from plumbing supply shops or hardware stores. The pipe selected should be 19 mm in diameter or larger. Some people report using up to 50 mm diameter pipe for this application. The surface area of the hollow pipe is greater than that of a solid rod of the same diameter. Because of certain current flow geometries in the system, however, the ground resistance is not half the resistance of a rod the same diameter, but is nonetheless lower.

Unlike the copper-clad steel rod, the copper pipe has no compression strength and will deform when hit with a hammer or other driving tool.

To overcome this problem (see Figure 13.2) you can use a garden hose as a 'water drill' to sink the ground rod.

Sweat solder a 'tee'-joint on the top end of the pipe, and then sweat solder a faucet or spigot fitting that matches the garden hose end on one end of the 'tee'-joint. Cap off the other port of the 'tee'-joint. Use the 'tee'-joint as a handle to drive the pipe into the ground. When water pressure is applied, the pipe will sink slowly into the ground as you apply a downward pressure on the tee-handle. In some cases, the pipe will slip into the ground easily, requiring only a few minutes. In other cases, where the soil is hard or has a heavy clay content, it will take considerably greater effort and more time. When you finish the task, turn off the water and remove the garden hose. Some people also recommend unsoldering and removing the 'tee'-joint.

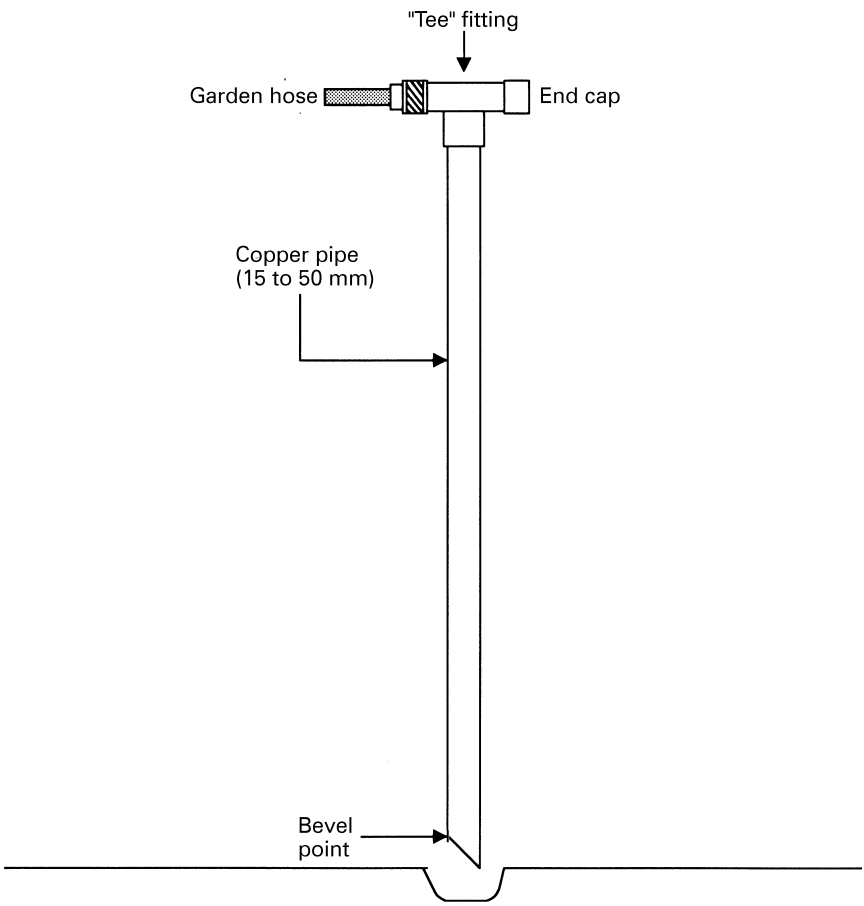


FIGURE 13.2

ALTERING SOIL CONDUCTIVITY

The conductivity of the soil determines how well, or how poorly, it conducts electrical current (Table 13.1). Moist soil over a brackish water dome conducts best. In the USA, the southern swamps make better radio station locations than most of the rest of the country; and the sand of the western deserts make the worst.

In Arizona, for example, one finds that few buildings have cellers or basements. The reason is that there is a thick calcium-based layer (called 'caleche') a meter or two down that resists even mechanized diggers. To build homes in that sand they have to drill holes in the caleche and then sink pylons down to bedrock.

The same caleche also prevents water from sinking into the sand, so it flows off the land in natural channels in the desert. The area is rather dangerous for out-of-state campers who do not understand that those nice flat, areas at the bottom of gulleys have a habit of filling in seconds if there is a distant thunder storm in the mountains. They look up, and to their horror see a 3–5 m high wall of water rushing toward them at 50 km/h! During the late summer months, Arizona has a 'monsoon' season when it rains hard and floods every day (90% of the state's annual rainfall comes in a 6-week period from sometime in July to mid-September). Evaporation is so rapid that an hour after the rain stops, there is little evidence of the rainfall except in the gulleys. If you have never felt hot rain, then visit southern Arizona in August – but be prepared for temperatures around 43°C in the shade and 80°C on the front seat of an automobile if you are so unwise as to leave all the windows rolled up.

Figure 13.3 shows a method for reducing the soil electrical resistance by treating with one of three chemicals: copper sulfate, magnesium sulfate or common rock salt. Rock salt is one of several salt materials used for snow and ice melting in snow-prone areas. If you cannot locate rock salt in a hardware store, then look for a store that sells ice cream making supplies. Rock salt is a principal ingredient in the process (but not the product).

TABLE 13.1

Type of soil	Dielectric constant	Conductivity (siemans/m)	Relative quality
Salt water	81	5	Best
Fresh water	80	0.001	Very poor
Pastoral hills	14–20	0.03–0.01	Very good
Marshy, wooded	12	0.0075	Average/poor
Rocky hills	12–14	10	Poor
Sandy	10	0.002	Poor
Cities	3–5	0.001	Very poor

Figure 13.3A shows a side sectioned view of a circular slit trench method of applying the chemical treatment to reduce ground resistance. **WARNING**: before using this method, you need to determine whether or not you really want to do it. The method involves putting conductive chemicals into the ground. *Make sure that local, state or national environmental regulations do not prohibit the practice*. Also, if you ever want to use the spot to grow vegetables or flowers – or even grass – it will prove difficult because the chemicals essentially poison the soil for several years. Fortunately, they are water-soluble, so the effect will not last forever.

Dig a 15–30 cm deep circular trench, about 30–60 cm from the ground rod. Fill that trench with a 10–12 cm layer of rock salt, magnesium sulfate, or copper sulfate (use only one). Water the trench well for about 15 minutes. Cover the remaining depth with soil removed when you have dug the trench. If you use about 100 kg of material, the treatment will need to be repeated every 24–36 months, depending upon the local rainfall and soil composition.

Figure 13.3B shows an alternative method: the saltpipe. Use either copper or PVC plumbing pipe, up to 10 cm in diameter (although 4–8 cm is easier to work). The overall length of the pipe should be at least 50–150 cm. Although longer pipes are useful, they are also difficult to install. Drill a large number of small holes in the pipe (no hole over 10 mm), sparing only the end that will be above the surface. Cap off both ends.

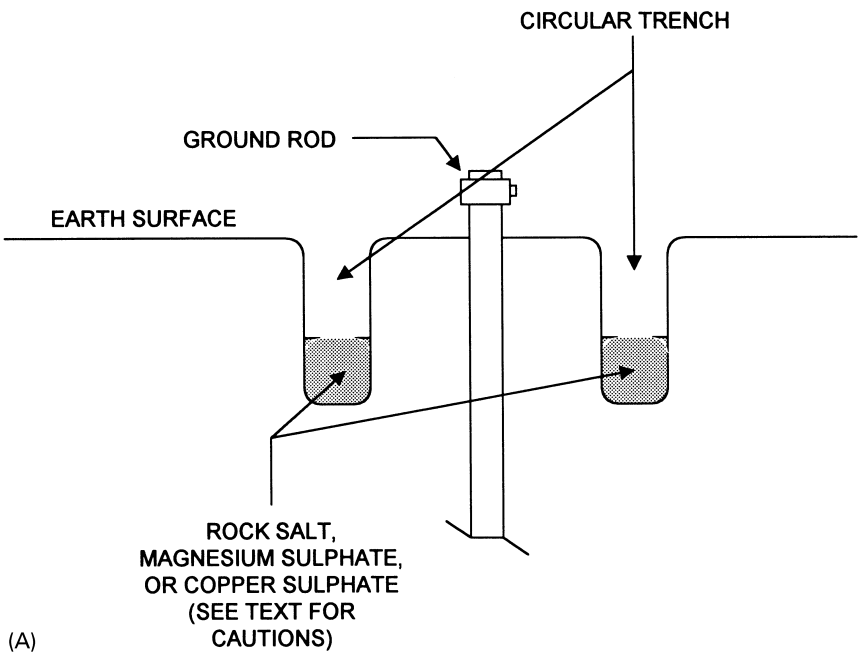


FIGURE 13.3

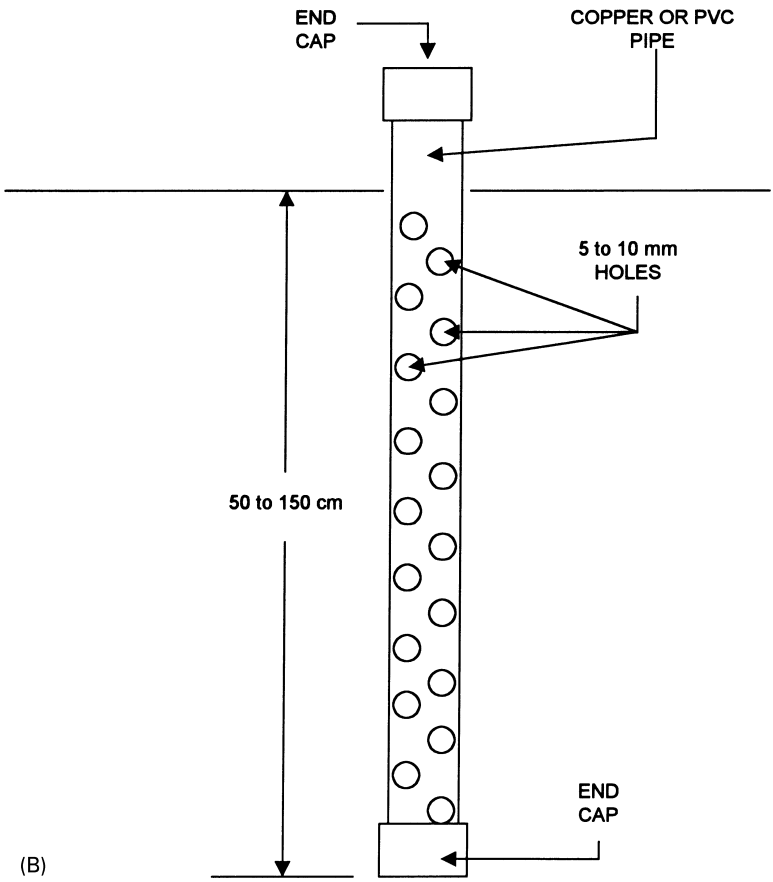


FIGURE 13.3 (*continued*)

Install several saltpipes in a pattern around the ground rod. Installation is best accomplished using a fencepost hole digger. Drop the saltpipe into the hole and backfill with water. Remove the top endcap and hose down for about 15 minutes, or until all the material is gone. Refill the pipe occasionally to account for the salt leaching out. When the chemical is completely leeched out of the pipe, refill the pipe and recap the top end. Leave the pipe in place.

Be certain of local environmental laws before using the above methods.

MULTIPLE GROUND RODS

The key to a low-resistance ground is the surface area in contact with the soil. One means for gaining surface area, and thereby reducing ground

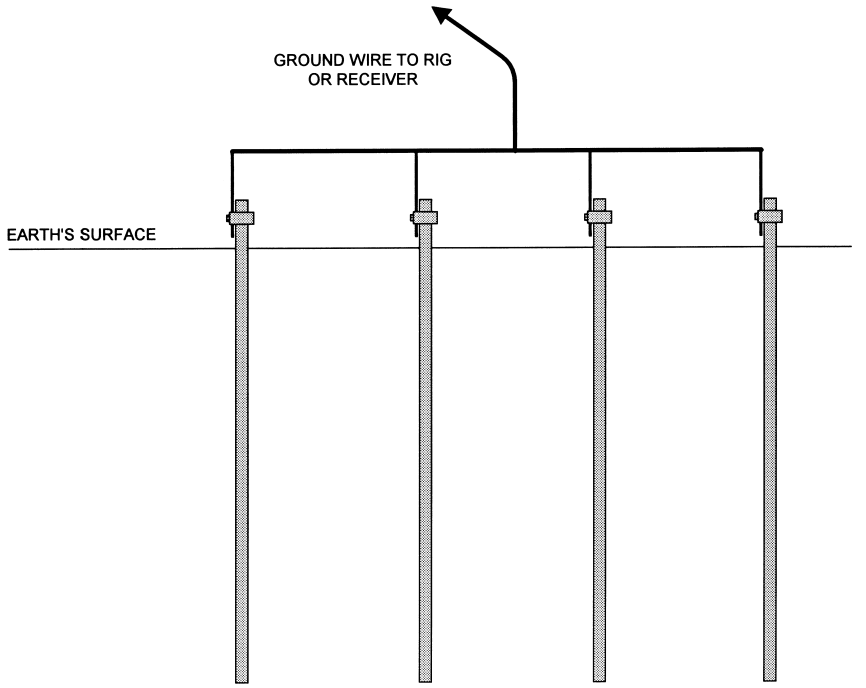


FIGURE 13.4

resistance, is to use multiple ground rods. Figure 13.4 shows the use of four ground rods in the same system. The rods are placed 30–100 cm apart for low- and medium-power levels, and perhaps more than 100 cm for higher amateur radio power levels. An electrical connection is made between the rods on the surface using either copper stripping or copper braid. The connections are sweat soldered in the manner described above, with the feedpoint at the center rod.

RADIALS AND COUNTERPOISE GROUNDS

The effectiveness of the ground system is enhanced substantially by the use of radials either above ground, on the surface, or buried under the surface. Figure 13.5 shows a vertical antenna with a set of ground radials. It is not unreasonable to use both radials and a ground rod. Vertical antennas are relatively ineffective unless provided with a good ground system, and for most installations that requirement is best met through a system of ground radials.

An effective system of radials requires a large number of radials. Although as few as two quarter-wavelength resonant radials will provide

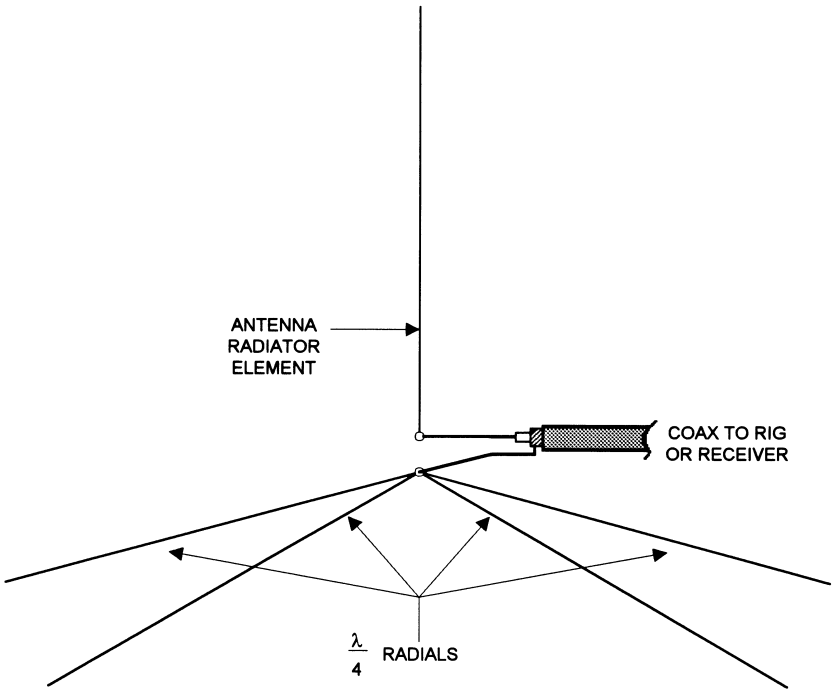


FIGURE 13.5

a significant improvement, the best performance is to use more. Broadcasters in the AM band (550–1640 kHz) are advised to use 120 half-wavelength radials. Installing more than 120 radials is both expensive and time-consuming, but does not provide any substantial improvement. For amateur and small commercial stations, a minimum of 16 quarter-wavelength radials should be used. Above ground, the use of insulated wire is recommended to reduce the possibility of anyone receiving RF burn if it is accidentally touched while transmitting. Although some sources claim that any size wire from AWG 26 up to AWG 10 (SWG 0–12) can be used, it is best to use larger sizes in that range (i.e. AWG 14 through AWG 10 (SWG 16–12)). Either solid or stranded wire can be used.

When viewed from above, the radials should be laid out in a symmetrical pattern around the antenna. This coverage provides both the lowest resistance and the best radiation pattern for the antenna. Solder all radials together at a common point, which might be the ground or mounting rod used to support the vertical antenna.

If your antenna is ground mounted, or if the radials are used with the station ground, then use a system such as that in Figure 13.6 to make the connections. A clamp on the ground rod is used to bind together the rod, the radials and the ground wire from the receiver or transmitter. The radials

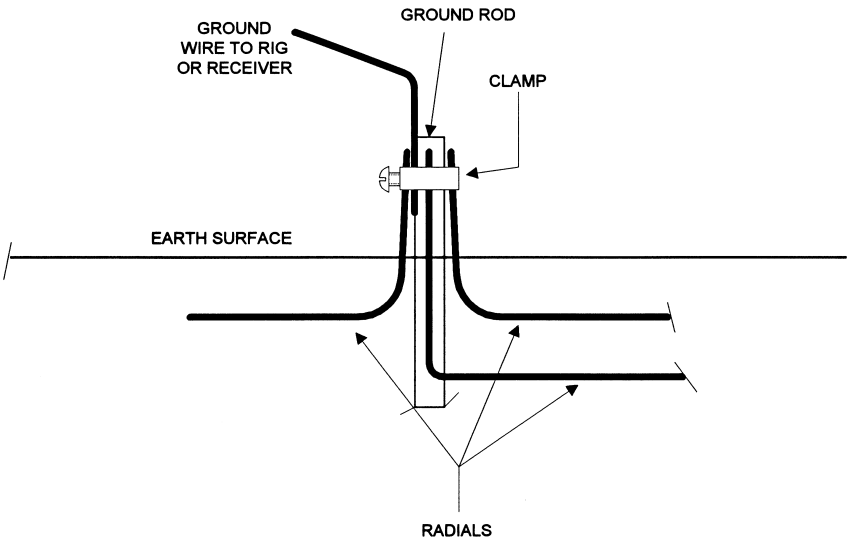


FIGURE 13.6

should be buried a few centimeters below the surface. Some people report using the tip of a spade or shovel to dig a very thin trench into which the wire can be pushed. Be careful to get the radial wire deep enough. Ordinary soil dynamics can make the wire rise to the surface within a year or so.

TUNING THE GROUND WIRE

An alternative that some operators use is the **ground wire tuner**. These instruments insert a series-tuned inductor–capacitor (LC network) in series with the ground line. These devices are usually called **artificial ground tuners**, and are inserted into the ground line of the station (Figure 13.7). Ground line tuners are especially recommended for cases where the wire to the ground is too long, in the case of single-wire antennas (e.g. Windom antennas or random-length wires), or anytime the tuning seems peculiar, but the antenna tuning unit works well into a dummy load. The purpose of the unit is to tune out any inductive or capacitive reactance on the line. Transmitter operators adjust the ground line tuner for maximum ground current at the operating frequency, although receiver operators will have to experiment a bit to find the correct settings. MFJ Enterprizes, Inc. (PO Box 494, Mississippi State, MS 39762, USA) makes several such devices. The version shown in Figure 13.8 is a stand-alone unit that contains only the ground current meter, inductor, and capacitor. The instrument in Figure 13.9 is a combination unit that includes both the ground line tuner and antenna tuner in one box.

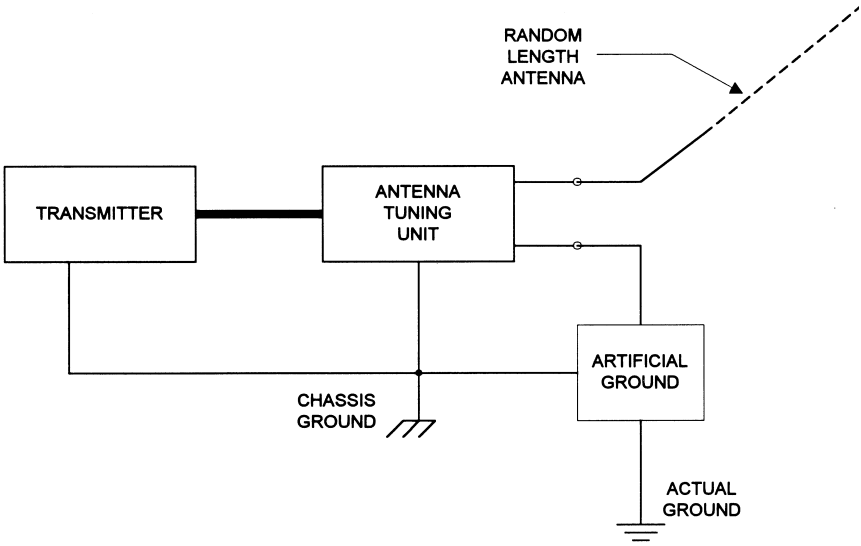


FIGURE 13.7

FIGURE 13.8

FIGURE 13.9

WARNING

Grounds for radio installations are intended for two purposes: making the radio antenna work better and lightning protection. Be absolutely sure to check with local authorities over the minimum requirements for grounding antenna structures – and then doing more than is minimally required. Not only is it possible to get into trouble with the authorities with a poor ground, you may also find that your homeowners and liability insurance is not valid if an uninspected installation is made. My own homeowners policy states that the insurance coverage is not in effect if an antenna is erected that meets the minimum requirements for a mechanical and electrical inspection by the county government but does not have a valid inspection certificate. Check with your local government and insurer before building any antenna or ground system.

CONCLUSION

Now that you have learned how to build safe and efficient antennas and grounds, in the words of Captain Piccard '*make it so*.'

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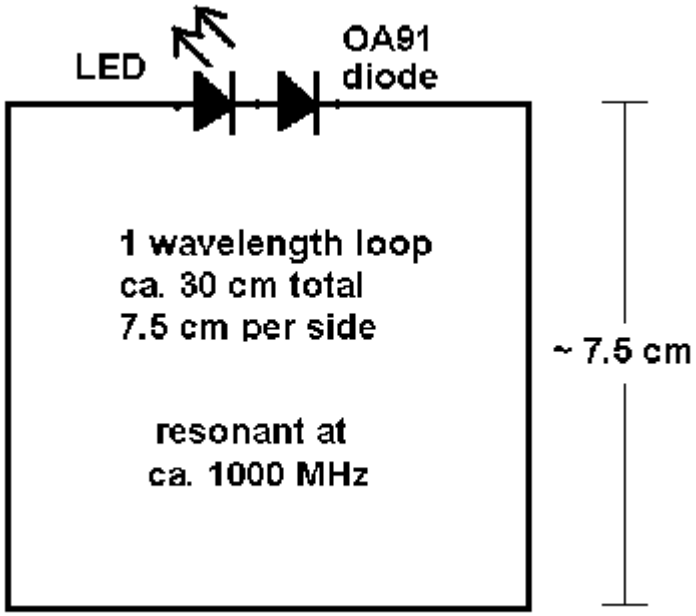
Simple demonstration to show mobile phones emit radio waves

Dr Jonathan Hare, Sussex University, Department of Physics, Falmer, Brighton. BN1 9QH
Note: this article is in press: Elektor Magazine, July-August 2010, p. 56-57

For other experiments with this device please see my full article at: mobile phone detector



left: mobile phone radio wave detector and right: the simple schematic. Below: detail of the LED and germanium diode.

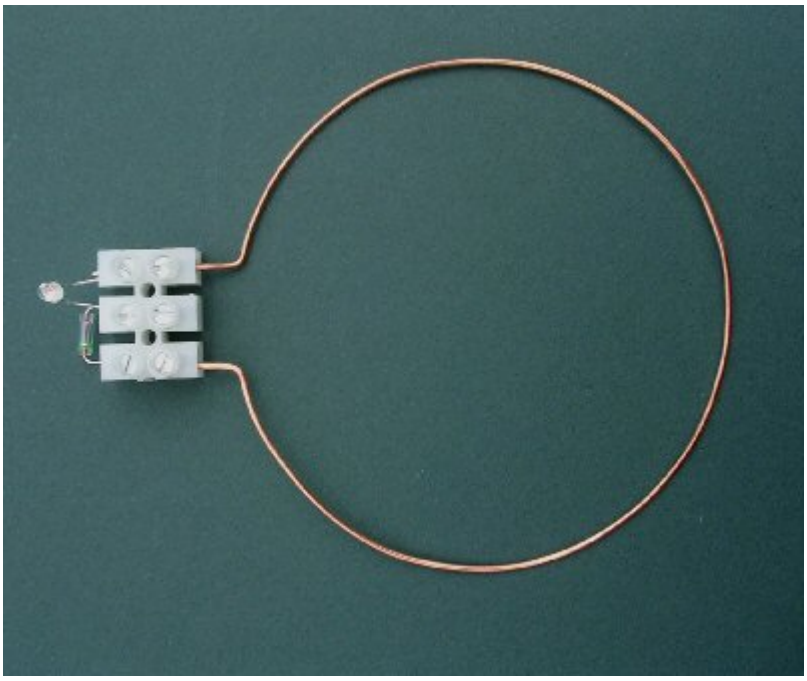
IMPORTANT NOTE: this device works very well on the old style mobile phones (as shown in the photo above). However, it does not always work well with modern smart phones. This may be because modern phones use higher frequencies, less power and use the power in a slightly different way (e.g. spread

spectrum). Some smart phones do work and success may be due to the signal strength of the local mobile phone mast nearby. If you are in a low signal area the phone will create more power to ensure reliable communications. If you are in a very strong signal area (very near the local network) your phone will drop its output power and consiquently there will be less power to pick-up and to convert to a voltage to light the LED.

This is a very simple and cheap device that demonstrates mobile phones ('cell phones' or 'handies') generate radio waves. We have a 30 cm (7.5 cm per side) full-wavelength loop antenna (a 'Quad' to radio amateurs) connected to a germanium diode and a hyper-bright LED. The loop can be made of copper wire, thin sheet metal or a track on a pcb. The diodes need to be wired correctly. I think the germanium diode is needed as the LED probably has too great a self-capacitance to perform at the very high AC frequencies generated by the phone (ca. 900 or 1800 Hz) but will work well with the DC pulses from the germanium diode (which has a very small capacitance).

To show the mobile generates radio waves put the mobile near to the loop and dial a number (use a free phone number, e.g. your voice mail) or text. The radio waves will induce a voltage into the loop, large enough to light the LED. The LED will flash indicating the digital data being sent by the mobile phone transmitter. You may need to set your phone to 'GSM 900/1800' rather than the '3G' network in the settings menu.

parts:
germanium diode: Maplin Electronics: QH71N or Rapid Electronics: 47-3114
LED: Maplin Electronics: UF72P or Rapid Electronics: 55-0085



A very simple connector block version and a circular 1 wavelength loop

THE CREATIVE SCIENCE CENTRE

Dr Jonathan Hare, Brighton, East Sussex. BN1 9QJ.

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NONRESIDENT TRAINING COURSE

October 1995

Electronics Technician

Volume 7—Antennas and Wave Propagation

NAVEDTRA 14092

DISTRIBUTION STATEMENT A: Approved for public release; distribution is unlimited.

Although the words "he," "him," and "his" are used sparingly in this course to enhance communication, they are not intended to be gender driven or to affront or discriminate against anyone.

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PREFACE

By enrolling in this self-study course, you have demonstrated a desire to improve yourself and the Navy. Remember, however, this self-study course is only one part of the total Navy training program. Practical experience, schools, selected reading, and your desire to succeed are also necessary to successfully round out a fully meaningful training program.

COURSE OVERVIEW: In completing this nonresident training course, you should be able to: discuss wave propagation in terms of the effects the earth's atmosphere has on it and the options available to receive optimum performance from equipment; identify communications and radar antennas using physical characteristics and installation location, radiation patterns, and power and frequency-handling capabilities. Be familiar with safety precautions for technicians working aloft; and discuss the different types of transmission lines in terms of physical structure, frequency limitations, electronic fields, and radiation losses.

THE COURSE: This self-study course is organized into subject matter areas, each containing learning objectives to help you determine what you should learn along with text and illustrations to help you understand the information. The subject matter reflects day-to-day requirements and experiences of personnel in the rating or skill area. It also reflects guidance provided by Enlisted Community Managers (ECMs) and other senior personnel, technical references, instructions, etc., and either the occupational or naval standards, which are listed in the *Manual of Navy Enlisted Manpower Personnel Classifications and Occupational Standards*, NAVPERS 18068.

THE QUESTIONS: The questions that appear in this course are designed to help you understand the material in the text.

VALUE: In completing this course, you will improve your military and professional knowledge. Importantly, it can also help you study for the Navy-wide advancement in rate examination. If you are studying and discover a reference in the text to another publication for further information, look it up.

1995 Edition Prepared by
ETC Larry D. Simmons
and
ETC Floyd L. Ace III

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Sailor's Creed

"I am a United States Sailor.

I will support and defend the Constitution of the United States of America and I will obey the orders of those appointed over me.

I represent the fighting spirit of the Navy and those who have gone before me to defend freedom and democracy around the world.

I proudly serve my country's Navy combat team with honor, courage and commitment.

I am committed to excellence and the fair treatment of all."

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2. Antennas 2-1

3. Introduction to Transmission and Waveguides 3-1

APPENDIX

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II. References AII-1

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SUMMARY OF THE ELECTRONICS TECHNICIAN TRAINING SERIES

This series of training manuals was developed to replace the *Electronics Technician 3 & 2* TRAMAN. The content is directed to personnel working toward advancement to Electronics Technician Second Class.

The nine volumes in the series are based on major topic areas with which the ET2 should be familiar. Volume 1, *Safety,* provides an introduction to general safety as it relates to the ET rating. It also provides both general and specific information on electronic tag-out procedures, man-aloft procedures, hazardous materials (i.e., solvents, batteries, and vacuum tubes), and radiation hazards. Volume 2, *Administration,* discusses COSAL updates, 3-M documentation, supply paperwork, and other associated administrative topics. Volume 3, *Communication Systems,* provides a basic introduction to shipboard and shore-based communication systems. Systems covered include man-pat radios (i.e., PRC-104, PSC-3) in the hf, vhf, uhf, SATCOM, and shf ranges. Also provided is an introduction to the Communications Link Interoperability System (CLIPS). Volume 4, *Radar Systems,* is a basic introduction to air search, surface search, ground controlled approach, and carrier controlled approach radar systems. Volume 5, *Navigation Systems, is* a basic introduction to navigation systems, such as OMEGA, SATNAV, TACAN, and man-pat systems. Volume 6, *Digital Data Systems, is* a basic introduction to digital data systems and includes discussions about SNAP II, laptop computers, and desktop computers. Volume 7, *Antennas and Wave Propagation,* is an introduction to wave propagation, as it pertains to Electronics Technicians, and shipboard and shore-based antennas. Volume 8, *Support Systems,* discusses system interfaces, troubleshooting, sub-systems, dry air, cooling, and power systems. Volume 9, *Electro-Optics, is* an introduction to night vision equipment, lasers, thermal imaging, and fiber optics.

INSTRUCTIONS FOR TAKING THE COURSE

ASSIGNMENTS

The text pages that you are to study are listed at the beginning of each assignment. Study these pages carefully before attempting to answer the questions. Pay close attention to tables and illustrations and read the learning objectives. The learning objectives state what you should be able to do after studying the material. Answering the questions correctly helps you accomplish the objectives.

SELECTING YOUR ANSWERS

Read each question carefully, then select the BEST answer. You may refer freely to the text. The answers must be the result of your own work and decisions. You are prohibited from referring to or copying the answers of others and from giving answers to anyone else taking the course.

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To have your assignments graded, you must be enrolled in the course with the Nonresident Training Course Administration Branch at the Naval Education and Training Professional Development and Technology Center (NETPDTC). Following enrollment, there are two ways of having your assignments graded: (1) use the Internet to submit your assignments as you complete them, or (2) send all the assignments at one time by mail to NETPDTC.

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We value your suggestions, questions, and criticisms on our courses. If you would like to communicate with us regarding this course, we encourage you, if possible, to use e-mail. If you write or fax, please use a copy of the Student Comment form that follows this page.

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NETPDTC 1550/41 (Rev 4-00

CHAPTER 1

WAVE PROPAGATION

The eyes and ears of a ship or shore station depend on sophisticated, highly computerized electronic systems. The one thing all of these systems have in common is that they lead to and from *antennas.* Ship's operators who must communicate, navigate, and be ready to fight the ship 24 hours a day depend on you to keep these emitters and sensors operational.

In this volume, we will review wave propagation, antenna characteristics, shore-based and shipboard communications antennas, matching networks, antenna tuning, radar antennas, antenna safety, transmission lines, connector installation and weatherproofing, waveguides, and waveguide couplings. When you have completed this chapter, you should be able to discuss the basic principles of wave propagation and the atmosphere's effects on wave propagation.

THE EARTH'S ATMOSPHERE

While radio waves traveling in free space have little outside influence to affect them, radio waves traveling in the earth's atmosphere have many influences that affect them. We have all experienced problems with radio waves, caused by certain atmospheric conditions complicating what at first seemed to be a relatively simple electronic problem. These problem-causing conditions result from a lack of uniformity in the earth's atmosphere.

Many factors can affect atmospheric conditions, either positively or negatively. Three of these are variations in geographic height, differences in geographic location, and changes in time (day, night, season, year).

To understand wave propagation, you must have at least a basic understanding of the earth's atmosphere. The earth's atmosphere is divided into three separate regions, or layers. They are the *troposphere,* the *stratosphere,* and the *ionosphere.* These layers are illustrated in figure 1-1.

TROPOSPHERE

Almost all weather phenomena take place in the troposphere. The temperature in this region decreases rapidly with altitude. Clouds form, and there may be a lot of turbulence because of variations in the temperature, pressure, and density. These conditions have a profound effect on the propagation of radio waves, as we will explain later in this chapter.

STRATOSPHERE

The stratosphere is located between the troposphere and the ionosphere. The temperature throughout this region is almost constant and there is little water vapor present. Because it is a relatively calm region with little or no temperature change, the stratosphere has almost no effect on radio waves.

IONOSPHERE

This is the most important region of the earth's atmosphere for long distance, point-to-point communications. Because the existence of the ionosphere is directly related to radiation emitted from the sun, the movement of the earth about the sun or changes in the sun's activity will result in variations in the ionosphere. These variations are of two general types: (1) those that more or less occur in cycles and, therefore, can be predicted with reasonable accuracy; and (2) those that are irregular as a result of abnormal behavior of the sun and, therefore, cannot be predicted. Both regular and irregular variations have important effects on radio-wave propagation. Since irregular variations cannot be predicted, we will concentrate on regular variations.

Regular Variations

The regular variations can be divided into four main classes: daily, 27-day, seasonal, and 11-year. We will concentrate our discussion on daily variations,

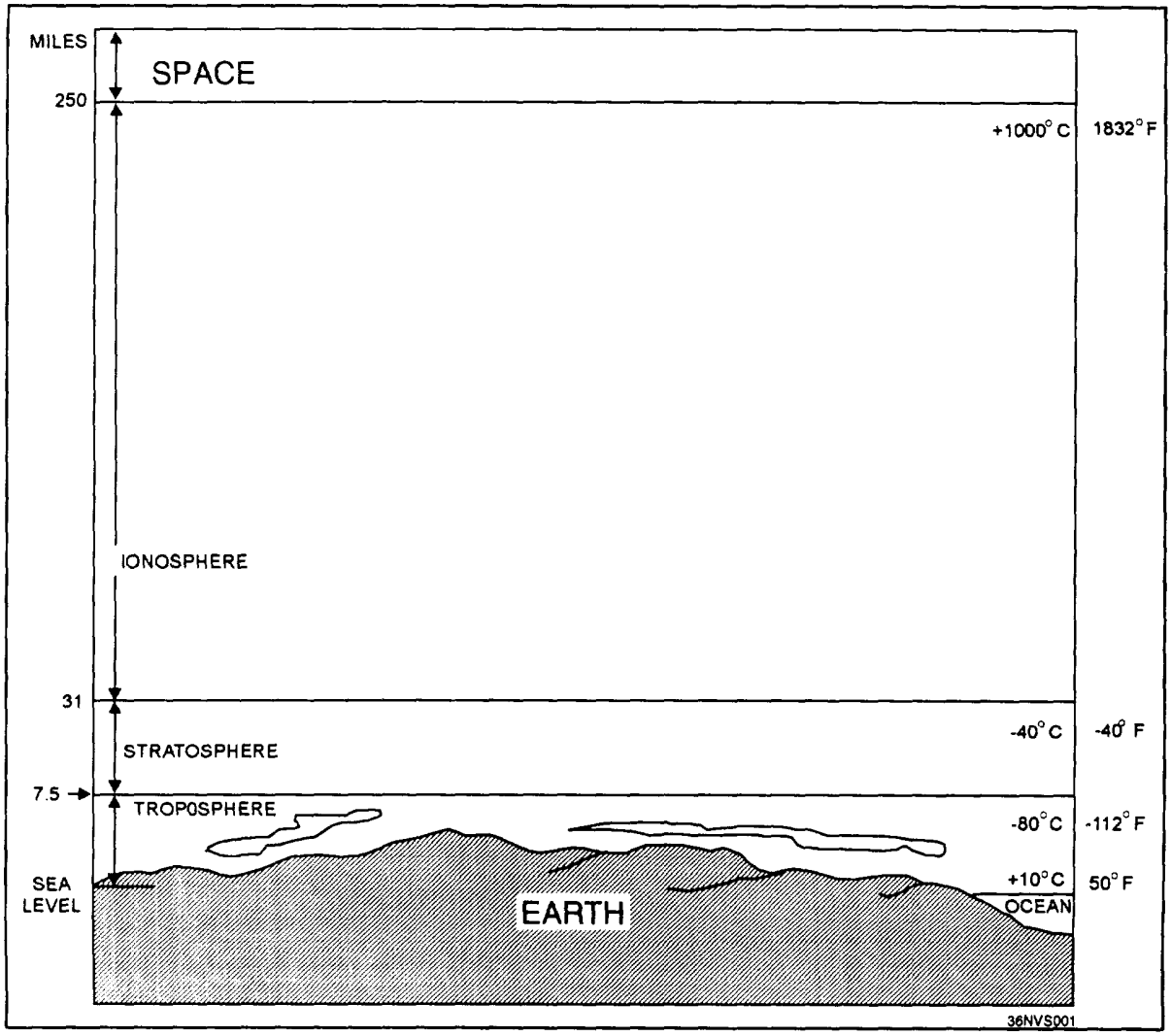


Figure 1.1—Atmospheric layers.

since they have the greatest effect on your job. Daily variations in the ionosphere produce four cloud-like layers of electrically-charged gas atoms called *ions*, which enable radio waves to be propagated great distances around the earth. Ions are formed by a process called *ionization*.

Ionization

In ionization, high-energy ultraviolet light waves from the sun periodically enter the ionosphere, strike neutral gas atoms, and knock one or more electrons free from each atom. When the electrons are knocked free, the atoms become positively charged **(positive ions)** and remain in space, along with the negatively-charged free electrons. The free electrons absorb some of the ultraviolet energy that initially set them free and form an ionized layer.

Since the atmosphere is bombarded by ultraviolet waves of differing frequencies, several ionized layers are formed at different altitudes. Ultraviolet waves of higher frequencies penetrate the most, so they produce ionized layers in the lower portion of the ionosphere. Conversely, ultraviolet waves of lower frequencies penetrate the least, so they form layers in the upper regions of the ionosphere.

An important factor in determining the density of these ionized layers is the elevation angle of the sun. Since this angle changes frequently, the height and thickness of the ionized layers vary, depending

on the time of day and the season of the year. Another important factor in determining layer density is known as *recombination*.

Recombination

Recombination is the reverse process of ionization. It occurs when free electrons and positive ions collide, combine, and return the positive ions to their original neutral state.

Like ionization, the recombination process depends on the time of day. Between early morning and late afternoon, the rate of ionization exceeds the rate of recombination. During this period the ionized layers reach their greatest density and exert maximum influence on radio waves. However, during the late afternoon and early evening, the rate of recombination exceeds the rate of ionization, causing the densities of the ionized layers to decrease. Throughout the night, density continues to decrease, reaching its lowest point just before sunrise. It is important to understand that this ionization and recombination process varies, depending on the ionospheric layer and the time of day. The following paragraphs provide an explanation of the four ionospheric layers.

Ionospheric Layers

The ionosphere is composed of three distinct layers, designated from lowest level to highest level **(D, E,** and **F)** as shown in figure 1-2. In addition, the **F** layer is divided into two layers, designated **F1** (the lower level) and **F2** (the higher level).

The presence or absence of these layers in the ionosphere and their height above the earth vary with the position of the sun. At high noon, radiation in the ionosphere above a given point is greatest, while at night it is minimum. When the radiation is removed, many of the particles that were ionized recombine. During the time between these two conditions, the position and number of ionized layers within the ionosphere change.

Since the position of the sun varies daily, monthly, and yearly with respect to a specific point on earth, the exact number of layers present is extremely difficult to determine. However, the following general statements about these layers can be made.

D LAYER.— The **D** layer ranges from about 30 to 55 miles above the earth. Ionization in the **D** layer is low because less ultraviolet light penetrates to this level. At very low frequencies, the **D** layer and the ground act as a huge waveguide, making communication possible only with large antennas and high-power transmitters. At low and medium frequencies, the **D** layer becomes highly absorptive, which limits the effective daytime communication range to about 200 miles. At frequencies above about 3 MHz, the **D** layer begins to lose its absorptive qualities.

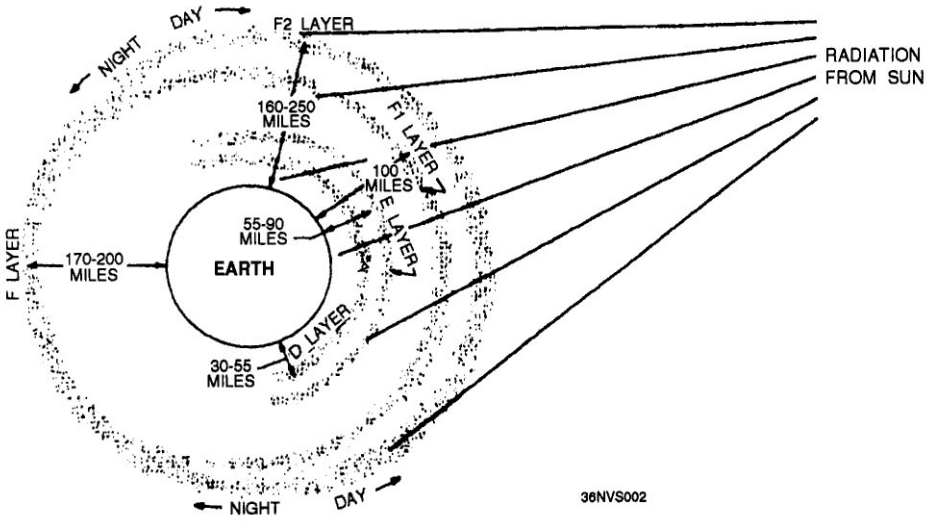


Figure 1-2.—Layers of the ionosphere.

Long-distance communication is possible at frequencies as high as 30 MHz. Waves at frequencies above this range pass through the **D** layer but are attenuated. After sunset. the **D** layer disappears because of the rapid recombination of ions. Low-frequency and medium-frequency long-distance communication becomes possible. This is why **AM** behaves so differently at night. Signals passing through the **D** layer normally are not absorbed but are propagated by the **E** and **F** layers.

E LAYER.— The **E** layer ranges from approximately 55 to 90 miles above the earth. The rate of ionospheric recombination in this layer is rather rapid after sunset, causing it to nearly disappear by midnight. The **E** layer permits medium-range communications on the low-frequency through very-high-frequency bands. At frequencies above about 150 MHz, radio waves pass through the **E** layer.

Sometimes a solar flare will cause this layer to ionize at night over specific areas. Propagation in this layer during this time is called SPORADIC-E. The range of communication in sporadic-E often exceeds 1000 miles, but the range is not as great as with F layer propagation.

F LAYER.— The **F** layer exists from about 90 to 240 miles above the earth. During daylight hours, the **F** layer separates into two layers, **F1** and **F2.** During the night, the **F1** layer usually disappears, The **F** layer produces maximum ionization during the afternoon hours, but the effects of the daily cycle are not as pronounced as in the **D** and **E** layers. Atoms in the **F** layer stay ionized for a longer time after sunset, and during maximum sunspot activity, they can stay ionized all night long.

Since the **F** layer is the highest of the ionospheric layers, it also has the longest propagation capability. For horizontal waves, the single-hop **F2** distance can reach 3000 miles. For signals to propagate over greater distances, multiple hops are required.

The **F** layer is responsible for most high-frequency, long-distance communications. The maximum frequency that the **F** layer will return depends on the degree of sunspot activity. During maximum sunspot activity, the **F** layer can return signals at frequencies as high as 100 MHz. During minimum sunspot activity, the maximum usable frequency can drop to as low as 10 MHz.

ATMOSPHERIC PROPAGATION

Within the atmosphere, radio waves can be refracted, reflected, and diffracted. In the following paragraphs, we will discuss these propagation characteristics.

REFRACTION

A radio wave transmitted into ionized layers is always refracted, or bent. This bending of radio waves is called *refraction*. Notice the radio wave shown in figure 1-3, traveling through the earth's atmosphere at a constant speed. As the wave enters the denser layer of charged ions, its upper portion moves faster than its lower portion. The abrupt speed increase of the upper part of the wave causes it to bend back toward the earth. This bending is always toward the propagation medium where the radio wave's velocity is the least.

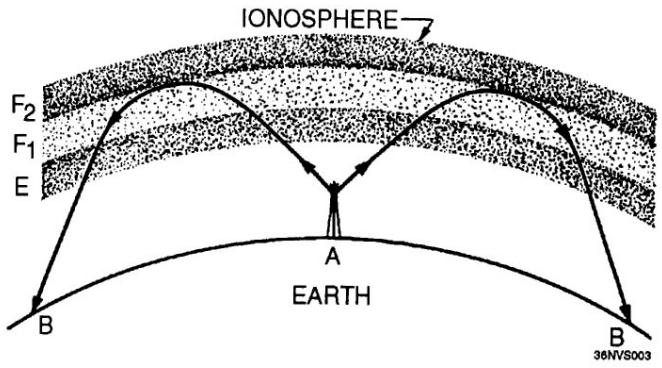


Figure 1-3.—Radio-wave refraction.

The amount of refraction a radio wave undergoes depends on three main factors.

1. The ionization density of the layer

2. The frequency of the radio wave

3. The angle at which the radio wave enters the layer

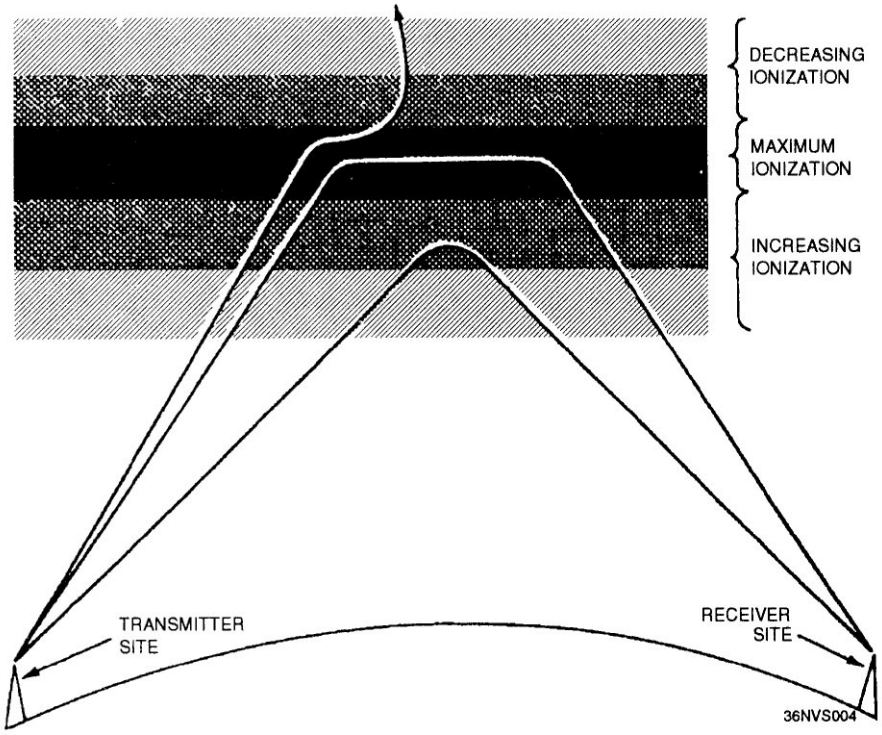


Figure 1-4.—Effects of ionospheric density on radio waves.

Layer Density

Figure 1-4 shows the relationship between radio waves and ionization density. Each ionized layer has a middle region of relatively dense ionization with less intensity above and below. As a radio wave enters a region of **increasing** ionization, a velocity increase causes it to bend back **toward** the earth. In the highly dense middle region, refraction occurs more slowly because the ionization density is uniform. As the wave enters the upper less dense region, the velocity of the upper part of the wave decreases and the wave is bent away from the earth.

Frequency

The lower the frequency of a radio wave, the more rapidly the wave is refracted by a given degree of ionization. Figure 1-5 shows three separate waves of differing frequencies entering the ionosphere at the same angle. You can see that the 5-MHz wave is refracted quite sharply, while the 20-MHz wave is refracted less sharply and returns to earth at a greater distance than the 5-MHz wave. Notice that the 100-MHz wave is lost into space. For any given ionized layer, there is a frequency, called the *escape point*, at which energy transmitted directly upward will escape into space. The maximum frequency just below the escape point is called the **critical frequency.** In this example, the 100-MHz wave's frequency is greater than the critical frequency for that ionized layer.

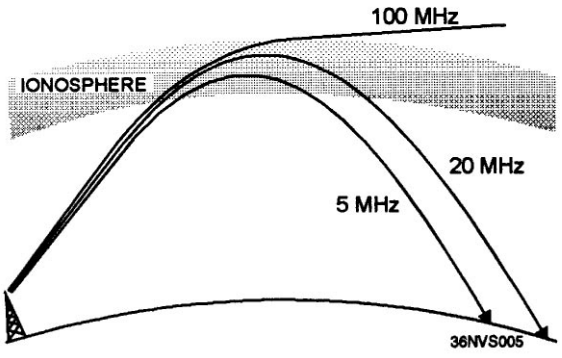


Figure 1-5.—Frequency versus refraction and distance.

The critical frequency of a layer depends upon the layer's density. If a wave passes through a

particular layer, it may still be refracted by a higher layer if its frequency is lower than the higher layer's critical frequency.

Angle of Incidence and Critical Angle

When a radio wave encounters a layer of the ionosphere, that wave is returned to earth at the same angle (roughly) as its *angle of incidence.* Figure 1-6 shows three radio waves of the same frequency entering a layer at different incidence angles. The angle at which wave A strikes the layer is too nearly vertical for the wave to be refracted to earth, However, wave B is refracted back to earth. The angle between wave B and the earth is called the **critical angle.** Any wave, at a given frequency, that leaves the antenna at an incidence angle greater than the critical angle will be lost into space. This is why wave A was not refracted. Wave C leaves the antenna at the smallest angle that will allow it to be refracted and still return to earth. The critical angle for radio waves depends on the layer density and the wavelength of the signal.

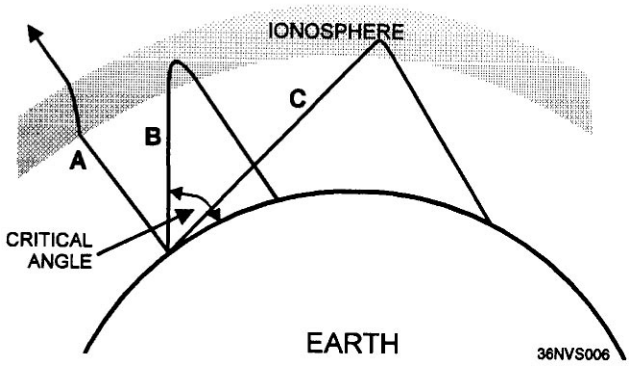


Figure 1-6.—Incidence angles of radio waves.

As the frequency of a radio wave is increased, the critical angle must be reduced for refraction to occur. Notice in figure 1-7 that the 2-MHz wave strikes the ionosphere at the critical angle for that frequency and is refracted. Although the 5-MHz line (broken line) strikes the ionosphere at a less critical angle, it still penetrates the layer and is lost As the angle is lowered, a critical angle is finally reached for the 5-MHz wave and it is refracted back to earth.

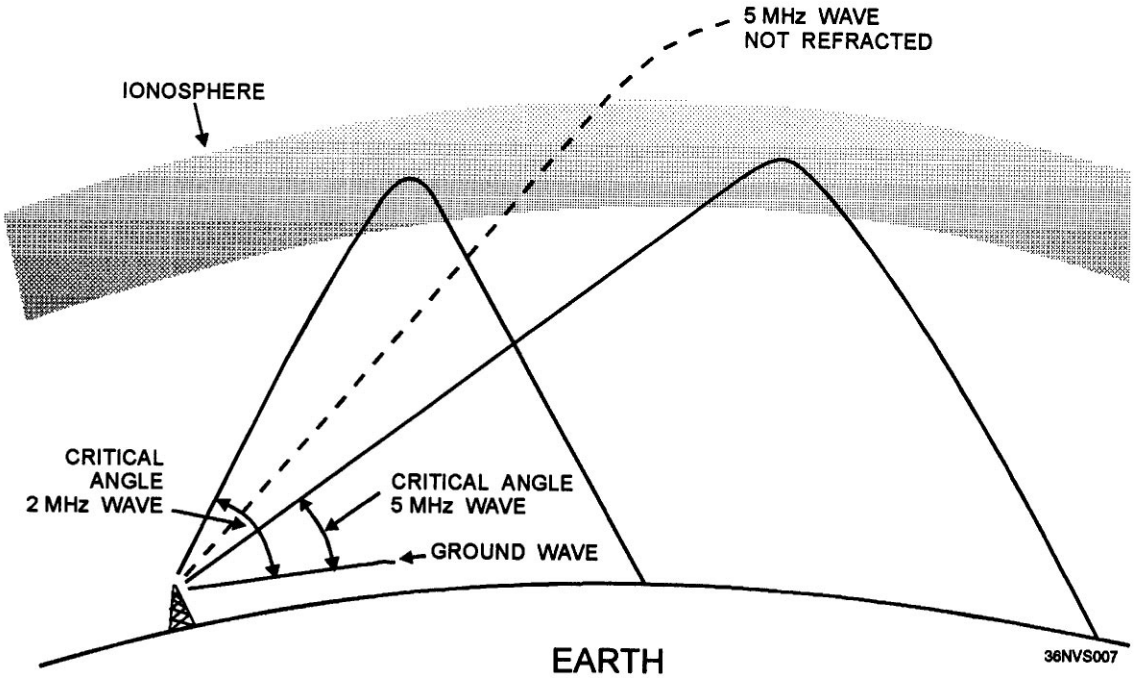


Figure 1-7.—Effect of frequency on the critical angle.

SKIP DISTANCE AND ZONE

Recall from your previous study that a transmitted radio wave separates into two parts, the sky wave and the ground wave. With those two components in mind, we will now briefly discuss *skip distance* and *skip zone.*

Skip Distance

Look at the relationship between the sky wave skip distance, skip zone, and ground wave coverage shown in figure 1-8. The *skip distance* is the distance from the transmitter to the point where the sky wave first returns to the earth. The skip distance depends on the wave's frequency and angle of incidence, and the degree of ionization.

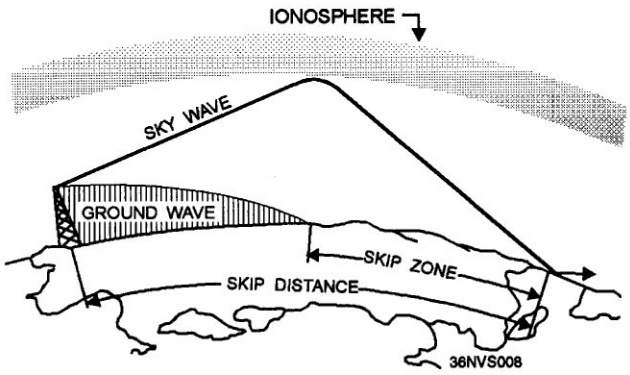


Figure 1-8.—Relationship between skip zone, skip distance, and ground wave.

Skip Zone

The *skip zone* is a zone of silence between the point where the ground wave is too weak for reception and the point where the sky wave is first returned to earth. The outer limit of the skip zone varies considerably, depending on the operating frequency, the time of day, the season of the year, sunspot activity, and the direction of transmission.

At very-low, low, and medium frequencies, a skip zone is never present. However, in the high-frequency spectrum, a skip zone is often present. As the operating frequency is increased, the skip zone widens to a point where the outer limit of the skip zone might be several thousand miles away. At frequencies above a certain maximum, the outer limit of the skip zone disappears completely, and no F-layer propagation is possible.

Occasionally, the first sky wave will return to earth within the range of the ground wave. In this case, severe fading can result from the phase difference between the two waves (the sky wave has a longer path to follow).

REFLECTION

Reflection occurs when radio waves are "bounced" from a flat surface. There are basically two types of reflection that occur in the atmosphere: earth reflection and ionospheric reflection. Figure 1-9 shows two

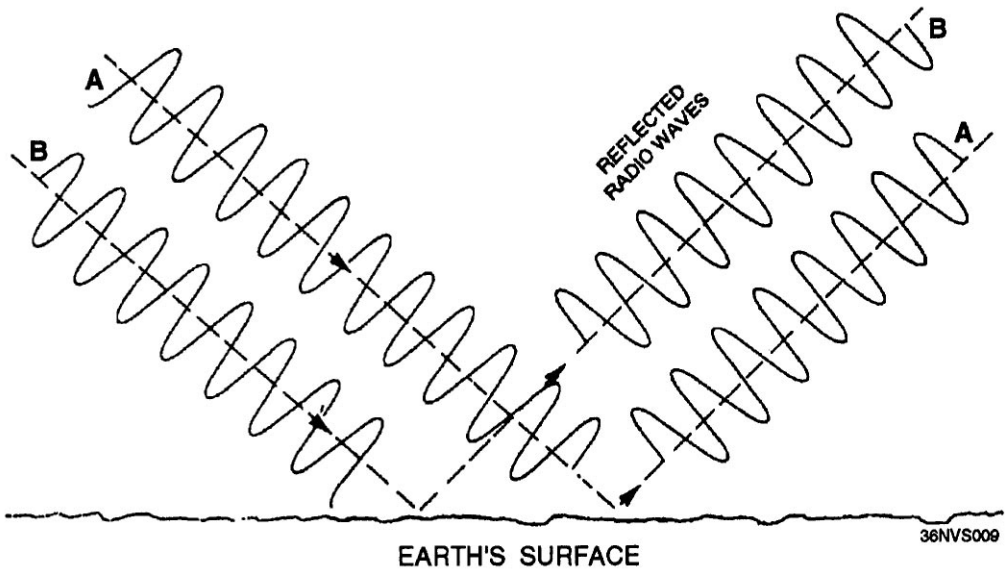


Figure 1-9.—Phase shift of reflected radio waves.

waves reflected from the earth's surface. Waves A and B bounce off the earth's surface like light off of a mirror. Notice that the positive and negative alternations of radio waves A and B are in phase before they strike the earth's surface. However, after reflection the radio waves are approximately 180 degrees out of phase. A phase shift has occurred.

The amount of phase shift that occurs is not constant. It varies, depending on the wave polarization and the angle at which the wave strikes the surface. Because reflection is not constant, fading occurs. Normally, radio waves reflected in phase produce stronger signals, while those reflected out of phase produce a weak or fading signal.

Ionospheric reflection occurs when certain radio waves strike a thin, highly ionized layer in the ionosphere. Although the radio waves are actually refracted, some may be bent back so rapidly that they appear to be reflected. For ionospheric reflection to occur, the highly ionized layer can be approximately no thicker than one wavelength of the wave. Since the ionized layers are often several miles thick, ionospheric reflection mostly occurs at long wavelengths (low frequencies).

DIFFRACTION

Diffraction is the ability of radio waves to turn sharp corners and bend around obstacles. Shown in figure 1-10, diffraction results in a change of direction of part of the radio-wave energy around the edges of an obstacle. Radio waves with long wavelengths compared to the diameter of an obstruction are easily propagated around the obstruction. However, as the wavelength decreases, the obstruction causes more and more attenuation, until at very-high frequencies a definite *shadow zone* develops. The shadow zone is basically a blank area on the opposite side of an obstruction in line-of-sight from the transmitter to the receiver.

Diffraction can extend the radio range beyond the horizon. By using high power and low-frequencies, radio waves can be made to encircle the earth by diffraction.

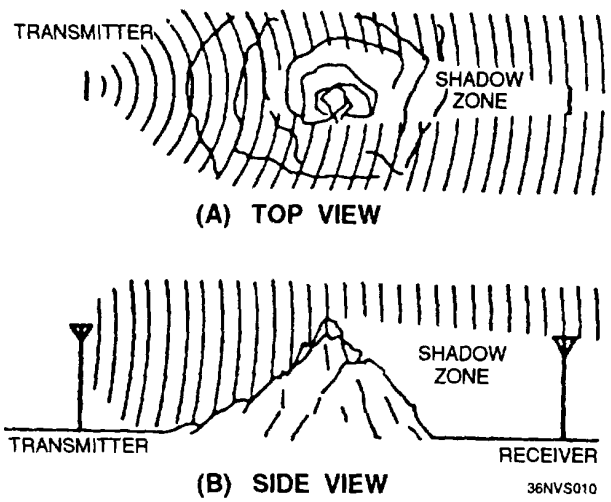


Figure 1-10.—Diffraction around an object.

ATMOSPHERIC EFFECTS ON PROPAGATION

As we stated earlier, changes in the ionosphere can produce dramatic changes in the ability to communicate. In some cases, communications distances are greatly extended. In other cases, communications distances are greatly reduced or eliminated. The paragraphs below explain the major problem of reduced communications because of the phenomena of fading and selective fading.

Fading

The most troublesome and frustrating problem in receiving radio signals is variations in signal strength, most commonly known as FADING. Several conditions can produce fading. When a radio wave is refracted by the ionosphere or reflected from the earth's surface, random changes in the polarization of the wave may occur. Vertically and horizontally mounted receiving antennas are designed to receive vertically and horizontally polarized waves, respectively. Therefore, changes in polarization cause changes in the received signal level because of the inability of the antenna to receive polarization changes.

Fading also results from absorption of the rf energy in the ionosphere. Most ionospheric absorption occurs in the lower regions of the ionosphere where ionization

density is the greatest. As a radio wave passes into the ionosphere, it loses some of its energy to the free electrons and ions present there. Since the amount of absorption of the radio-wave energy varies with the density of the ionospheric layers, there is no fixed relationship between distance and signal strength for ionospheric propagation. Absorption fading occurs for a longer period than other types of fading, since absorption takes place slowly. Under certain conditions, the absorption of energy is so great that communication over any distance beyond the line of sight becomes difficult.

Although fading because of absorption is the most serious type of fading, fading on the ionospheric circuits is mainly a result of multipath propagation.

Multipath Fading

MULTIPATH is simply a term used to describe the multiple paths a radio wave may follow between transmitter and receiver. Such propagation paths include the ground wave, ionospheric refraction, reradiation by the ionospheric layers, reflection from the earth's surface or from more than one ionospheric layer, and so on. Figure 1-11 shows a few of the paths that a signal can travel between two sites in a typical circuit. One path, XYZ, is the basic ground wave. Another path, XFZ, refracts the wave at the F layer and passes it on to the receiver at point Z. At point Z, the received signal is a combination of the ground wave and the sky wave. These two signals, having traveled different paths, arrive at point Z at different times. Thus, the arriving waves may or may not be in phase with each other. A similar situation may result at point A. Another path, XFZFA, results from a greater angle of incidence and two refractions from the F layer. A wave traveling that path and one traveling the XEA path may or may not arrive at point A in phase. Radio waves that are received in phase reinforce each other and produce a stronger signal at the receiving site, while those that are received out of phase produce a weak or fading signal. Small alterations in the transmission path may change the phase relationship of the two signals, causing periodic fading.

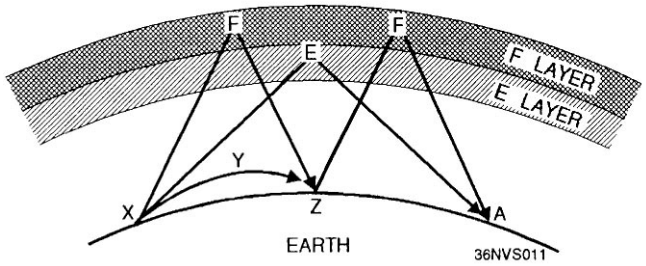


Figure 1-11.—Multipath transmission.

Multipath fading may be minimized by practices called SPACE DIVERSITY and FREQUENCY DIVERSITY In space diversity, two or more receiving antennas are spaced some distance apart. Fading does not occur simultaneously at both antennas. Therefore, enough output is almost always available from one of the antennas to provide a useful signal.

In frequency diversity, two transmitters and two receivers are used, each pair tuned to a different frequency, with the same information being transmitted simultaneously over both frequencies. One of the two receivers will almost always produce a useful signal.

Selective Fading

Fading resulting from multipath propagation varies with frequency since each frequency arrives at the
receiving point via a different radio path. When a wide band of frequencies is transmitted simultaneously,
each frequency will vary in the amount of fading. This variation is called SELECTIVE FADING. When selective fading occurs, all frequencies of the transmitted signal do not retain their original phases and relative amplitudes. This fading causes severe distortion of the signal and limits the total signal transmitted.

Frequency shifts and distance changes because of daily variations of the different ionospheric layers are summarized in table 1-1.

Table 1-1.–Daily Ionospheric Communications

D LAYER: reflects vlf waves for long-range communications; refracts lf and mf for short-range communications; has little effect on vhf and above; gone at night.
E LAYER: depends on the angle of the sun: refracts hf waves during the day up to 20 MHz to distances of 1200 miles: greatly reduced at night.
F LAYER: structure and density depend on the time of day and the angle of the sun: consists of one layer at night and splits into two layers during daylight hours.
F1 LAYER: density depends on the angle of the sun; its main effect is to absorb hf waves passing through to the F2 layer.
F2 LAYER: provides long-range hf communications; very variable; height and density change with time of day, season, and sunspot activity.

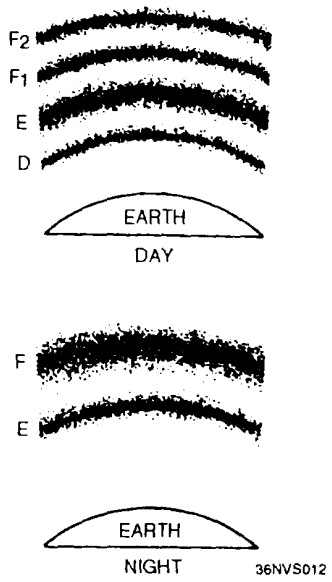


Figure 1-12.—Ionospheric layers.

OTHER PHENOMENA THAT AFFECT COMMUNICATIONS

Although daily changes in the ionosphere have the greatest effect on communications, other phenomena also affect communications, both positively and negatively. Those phenomena are discussed briefly in the following paragraphs.

SEASONAL VARIATIONS IN THE IONOSPHERE

Seasonal variations are the result of the earth's revolving around the sun, because the relative position of the sun moves from one hemisphere to the other with the changes in seasons. Seasonal variations of the D, E, and F1 layers are directly related to the highest angle of the sun, meaning the ionization density of these layers is greatest during the summer. The F2 layer is just the opposite. Its ionization is greatest during the winter, Therefore, operating frequencies for F2 layer propagation are higher in the winter than in the summer.

SUNSPOTS

One of the most notable occurrences on the surface of the sun is the appearance and disappearance of dark, irregularly shaped areas known as SUNSPOTS. Sunspots are believed to be caused by violent eruptions on the sun and are characterized by strong magnetic fields. These sunspots cause variations in the ionization level of the ionosphere.

Sunspots tend to appear in two cycles, every 27 days and every 11 years.

Twenty-Seven Day Cycle

The number of sunspots present at any one time is constantly changing as some disappear and new ones emerge. As the sun rotates on its own axis, these sunspots are visible at 27-day intervals, which is the approximate period for the sun to make one complete revolution. During this time period, the fluctuations in ionization are greatest in the F2 layer. For this reason, calculating critical frequencies for long-distance communications for the F2 layer is not possible and allowances for fluctuations must be made.

Eleven-Year Cycle

Sunspots can occur unexpectedly, and the life span of individual sunspots is variable. The ELEVEN-YEAR SUN SPOT CYCLE is a regular cycle of sunspot activity that has a minimum and maximum level of activity that occurs every 11 years. During periods of maximum activity, the ionization density of all the layers increases. Because of this, the absorption in the D layer increases and the critical frequencies for the E, F1, and F2 layers are higher. During these times, higher operating frequencies must be used for long-range communications.

IRREGULAR VARIATIONS

Irregular variations are just that, unpredictable changes in the ionosphere that can drastically affect our ability to communicate. The more common variations are sporadic E, ionospheric disturbances, and ionospheric storms.

Sporadic E

Irregular cloud-like patches of unusually high ionization, called the sporadic E, often format heights near the normal E layer. Their exact cause is not known and their occurrence cannot be predicted. However, sporadic E is known to vary significantly with latitude. In the northern latitudes, it appears to be closely related to the aurora borealis or northern lights.

The sporadic E layer can be so thin that radio waves penetrate it easily and are returned to earth by the upper layers, or it can be heavily ionized and extend up to several hundred miles into the ionosphere. This condition may be either harmful or helpful to radio-wave propagation.

On the harmful side, sporadic E may blank out the use of higher more favorable layers or cause additional absorption of radio waves at some frequencies. It can also cause additional multipath problems and delay the arrival times of the rays of RF energy.

On the helpful side, the critical frequency of the sporadic E can be greater than double the critical frequency of the normal ionospheric layers. This may permit long-distance communications with unusually high frequencies. It may also permit short-distance communications to locations that would normally be in the skip zone.

Sporadic E can appear and disappear in a short time during the day or night and usually does not occur at same time for all transmitting or receiving stations.

Sudden Ionospheric Disturbances

Commonly known as SID, these disturbances may occur without warning and may last for a few minutes to several hours. When SID occurs, long-range hf communications are almost totally blanked out. The radio operator listening during this time will believe his or her receiver has gone dead.

The occurrence of SID is caused by a bright solar eruption producing an unusually intense burst of ultraviolet light that is not absorbed by the F1, F2, or E layers. Instead, it causes the D-layer ionization density to greatly increase. As a result, frequencies above 1 or 2 megahertz are unable to penetrate the D layer and are completely absorbed.

Ionospheric Storms

Ionospheric storms are caused by disturbances in the earth's magnetic field. They are associated with both solar eruptions and the 27-day cycle, meaning they are related to the rotation of the sun. The effects of ionospheric storms are a turbulent ionosphere and very erratic sky-wave propagation. The storms affect mostly the F2 layer, reducing its ion density and causing the critical frequencies to be lower than

normal. What this means for communication purposes is that the range of frequencies on a given circuit is smaller than normal and that communications are possible only at lower working frequencies.

Weather

Wind, air temperature, and water content of the atmosphere can combine either to extend radio communications or to greatly attenuate wave propagation. making normal communications extremely difficult. Precipitation in the atmosphere has its greatest effect on the higher frequency ranges. Frequencies in the hf range and below show little effect from this condition.

RAIN.— Attenuation because of raindrops is greater than attenuation for any other form of precipitation. Raindrop attenuation may be caused either by absorption, where the raindrop acts as a poor dielectric, absorbs power from the radio wave and dissipates the power by heat loss; or by scattering (fig. 1-13). Raindrops cause greater attenuation by scattering than by absorption at frequencies above 100 megahertz. At frequencies above 6 gigahertz, attenuation by raindrop scatter is even greater.

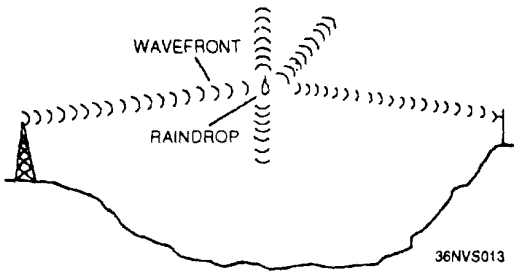


Figure 1-13.–Rf energy losses from scattering.

FOG.— Since fog remains suspended in the atmosphere, the attenuation is determined by the quantity of water per unit volume (density of the fog) and by the size of the droplets. Attenuation because of fog has little effect on frequencies lower than 2 gigahertz, but can cause serious attenuation by absorption at frequencies above 2 gigahertz.

SNOW.— Since snow has about 1/8 the density of rain, and because of the irregular shape of the snowflake, the scattering and absorption losses are difficult to compute, but will be less than those caused by raindrops.

HAIL.— Attenuation by hail is determined by the size of the stones and their density. Attenuation of radio waves by scattering because of hailstones is considerably less than by rain.

TEMPERATURE INVERSION

When layers of warm air form above layers of cold air, the condition known as temperature inversion develops. This phenomenon causes ducts or channels to be formed, by sandwiching cool air either between the surface of the earth and a layer of warm air, or between two layers of warm air. If a transmitting antenna extends into such a duct, or if the radio wave enters the duct at a very low angle of incidence, vhf and uhf transmissions may be propagated far beyond normal line-of-sight distances. These long distances are possible because of the different densities and refractive qualities of warm and cool air. The sudden change in densities when a radio wave enters the warm air above the duct causes the wave to be refracted back toward earth. When the wave strikes the earth or a warm layer below the duct, it is again reflected or refracted upward and proceeds on through the duct with a multiple-hop type of action. An example of radio-wave propagation by ducting is shown in figure 1-14.

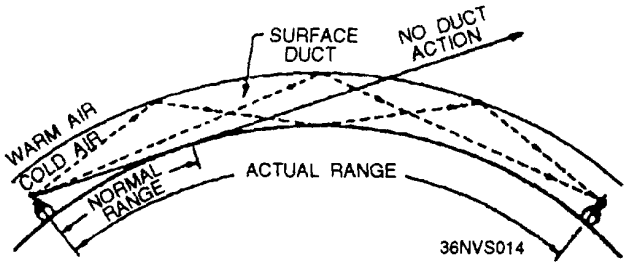


Figure 1-14.—Duct effect caused by temperature inversion.

TRANSMISSION LOSSES

All radio waves propagated over the ionosphere undergo energy losses before arriving at the receiving site. As we discussed earlier, absorption and lower

atmospheric levels in the ionosphere account for a large part of these energy losses. There are two other types of losses that also significantly affect propagation. These losses are known as *ground reflection losses* and *freespace loss.* The combined effect of absorption ground reflection loss, and freespace loss account for most of the losses of radio transmissions propagated in the ionosphere.

GROUND REFLECTION LOSS

When propagation is accomplished via multihop refraction, rf energy is lost each time the radio wave is reflected from the earth's surface. The amount of energy lost depends on the frequency of the wave, the angle of incidence, ground irregularities, and the electrical conductivity of the point of reflection.

FREESPACE LOSS

Normally, the major loss of energy is because of the spreading out of the wavefront as it travels from the transmitter. As distance increases, the area of the wavefront spreads out, much like the beam of a flashlight. This means the amount of energy contained within any unit of area on the wavefront decreases as distance increases. By the time the energy arrives at the receiving antenna, the wavefront is so spread out that the receiving antenna extends into only a small portion of the wavefront. This is illustrated in figure 1-15.

FREQUENCY SELECTION

You must have a thorough knowledge of radio-wave propagation to exercise good judgment when selecting transmitting and receiving antennas and operating frequencies. Selecting a usable operating frequency within your given allocations and availability is of prime importance to maintaining reliable communications.

For successful communication between any two specified locations at any given time of the day, there is a maximum frequency, a lowest frequency and an optimum frequency that can be used.

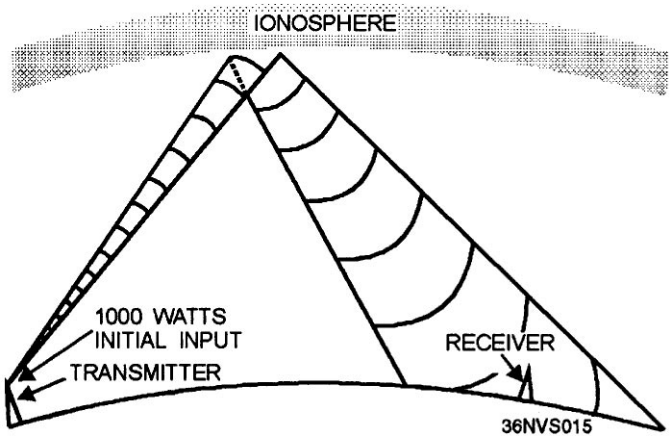


Figure 1-15.—Freespace loss principle.

MAXIMUM USABLE FREQUENCY

The higher the frequency of a radio wave, the lower the rate of refraction by the ionosphere. Therefore, for a given angle of incidence and time of day, there is a maximum frequency that can be used for communications between two given locations. This frequency is known as the **MAXIMUM USABLE FREQUENCY** (muf).

Waves at frequencies above the muf are normally refracted so slowly that they return to earth beyond the desired location or pass on through the ionosphere and are lost. Variations in the ionosphere that can raise or lower a predetermined muf may occur at anytime. his is especially true for the highly variable F2 layer.

LOWEST USABLE FREQUENCY

Just as there is a muf that can be used for communications between two points, there is also a minimum operating frequency that can be used known as the LOWEST USABLE FREQUENCY (luf). As the frequency of a radio wave is lowered, the rate of refraction increases. So a wave whose frequency is below the established luf is refracted back to earth at a shorter distance than desired, as shown in figure 1-16.

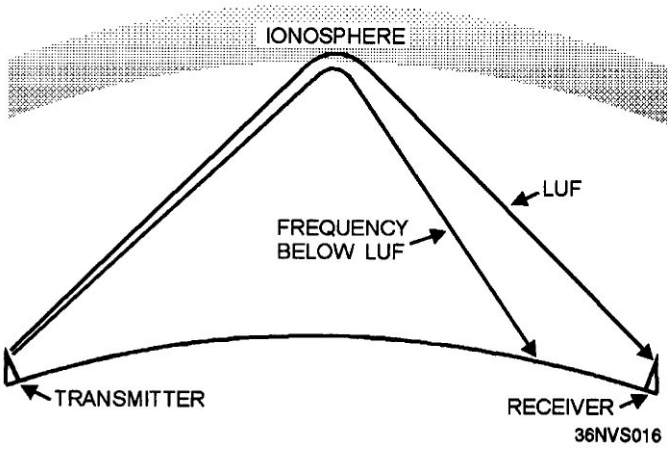


Figure 1-16.—Refraction of frequencies below the lowest usable frequency (luf).

As a frequency is lowered, absorption of the radio wave increases. A wave whose frequency is too low is absorbed to such an extent that it is too weak for reception. Atmospheric noise is also greater at lower frequencies. A combination of higher absorption and atmospheric noise could result in an unacceptable signal-to-noise ratio.

For a given angle ionospheric conditions, of incidence and set of the luf depends on the refraction properties of the ionosphere, absorption considerations, and the amount of noise present.

OPTIMUM WORKING FREQUENCY

The most practical operating frequency is one that you can rely onto have the least number of problems. It should be high enough to avoid the problems of multipath fading, absorption, and noise encountered at the lower frequencies; but not so high as to be affected by the adverse effects of rapid changes in the ionosphere.

A frequency that meets the above criteria is known as the OPTIMUM WORKING FREQUENCY It is abbreviated “fot” from the initial letters of the French words for optimum working frequency, “frequence optimum de travail.” The fot is roughly about 85% of the muf, but the actual percentage varies and may be considerably more or less than 85 percent.

In this chapter, we discussed the basics of radio-wave propagation and how atmospheric conditions determine the operating parameters needed to ensure successful communications. In chapter 2, we will discuss basic antenna operation and design to complete your understanding of radio-wave propagation.

CHAPTER 2

ANTENNAS

As an Electronics Technician, you are responsible for maintaining systems that both radiate and receive electromagnetic energy. Each of these systems requires some type of antenna to make use of this electromagnetic energy. In this chapter we will discuss antenna characteristics, different antenna types, antenna tuning, and antenna safety.

ANTENNA CHARACTERISTICS

An antenna may be defined as a conductor or group of conductors used either for radiating electromagnetic energy into space or for collecting it from space. Electrical energy from the transmitter is converted into electromagnetic energy by the antenna and radiated into space. On the receiving end, electromagnetic energy is converted into electrical energy by the antenna and fed into the receiver.

The electromagnetic radiation from an antenna is made up of two components, the E field and the H field. The total energy in the radiated wave remains constant in space except for some absorption of energy by the earth. However, as the wave advances, the energy spreads out over a greater area. This causes the amount of energy in a given area to decrease as distance from the source increases.

The design of the antenna system is very important in a transmitting station. The antenna must be able to radiate efficiently so the power supplied by the transmitter is not wasted. An efficient transmitting antenna must have exact dimensions, determined by the frequency being transmitted. The dimensions of the receiving antenna are not critical for relatively low frequencies, but their importance increases drastically as the transmitted frequency increases.

Most practical transmitting antennas are divided into two basic classifications, HERTZ ANTENNAS (half-wave) and MARCONI (quarter-wave) ANTENNAS. Hertz antennas are generally installed some distance above the ground and are positioned to radiate either vertically or horizontally. Marconi antennas operate with one end grounded and are mounted perpendicular to the earth or a surface acting as a ground. The Hertz antenna, also referred to as a dipole, is the basis for some of the more complex antenna systems used today. Hertz antennas are generally used for operating frequencies of 2 MHz and above, while Marconi antennas are used for operating frequencies below 2 MHz.

All antennas, regardless of their shape or size, have four basic characteristics: reciprocity, directivity, gain, and polarization.

RECIPROCITY

RECIPROCITY is the ability to use the same antenna for both transmitting and receiving. The electrical characteristics of an antenna apply equally, regardless of whether you use the antenna for transmitting or receiving. The more efficient an antenna is for transmitting a certain frequency, the more efficient it will be as a receiving antenna for the same frequency. This is illustrated by figure 2-1, view A. When the antenna is used for transmitting, maximum radiation occurs at right angles to its axis. When the same antenna is used for receiving (view B), its best reception is along the same path; that is, at right angles to the axis of the antenna.

DIRECTIVITY

The DIRECTIVITY of an antenna or array is a measure of the antenna's ability to focus the energy in one or more specific directions. You can determine an antenna's directivity by looking at its radiation pattern. In an array propagating a given amount of energy, more radiation takes place in certain directions than in others. The elements in the array can be arranged so they change the pattern and distribute the energy more evenly in all directions. The opposite is also possible. The elements can be arranged so the radiated energy is *focused* in one direction. The

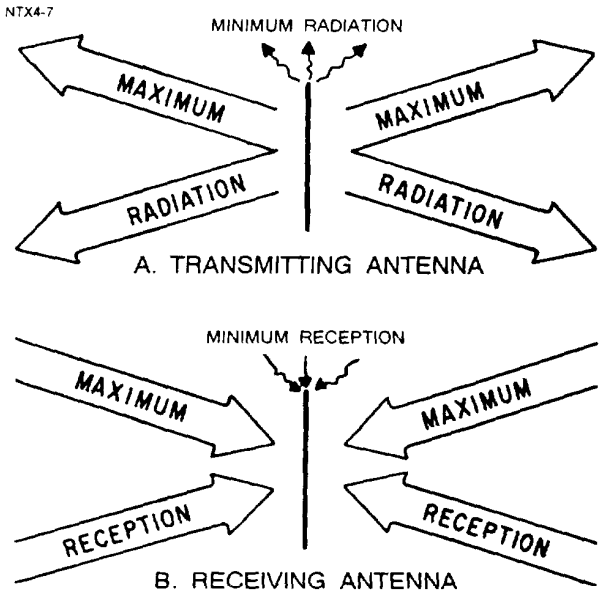


Figure 2-1.—Reciprocity of antennas.

elements can be considered as a group of antennas fed from a common source.

GAIN

As we mentioned earlier, some antennas are highly directional. That is, they propagate more energy in certain directions than in others. The ratio between the amount of energy propagated in these directions and the energy that would be propagated if the antenna were not directional is known as antenna GAIN. The gain of an antenna is constant. whether the antenna is used for transmitting or receiving.

POLARIZATION

Energy from an antenna is radiated in the form of an expanding sphere. A small section of this sphere is called a wavefront. positioned perpendicular to the direction of the radiation field (fig. 2-2). Within this wavefront. all energy is in phase. Usually, all points on the wavefront are an equal distance from the antenna. The farther from the antenna the wave is, the less curved it appears. At a considerable distance, the wavefront can be considered as a plane surface at right angles to the direction of propagation.

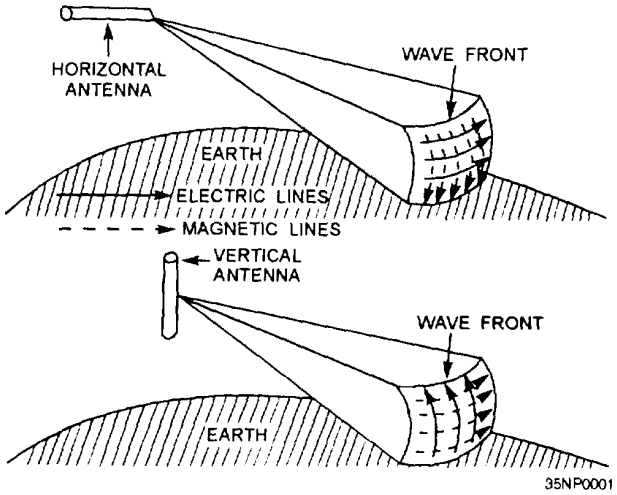


Figure 2-2.—Horizontal and vertical polarization.

The radiation field is made up of magnetic and electric lines of force that are always at right angles to each other. Most electromagnetic fields in space are said to be linearly polarized. The direction of polarization is the direction of the electric vector. That is, if the electric lines of force (E lines) are horizontal, the wave is said to be horizontally polarized (fig. 2-2), and if the E lines are vertical, the wave is said to be vertically polarized. Since the electric field is parallel to the axis of the dipole, the antenna is in the plane of polarization.

A horizontally placed antenna produces a horizontally polarized wave, and a vertically placed antenna produces a vertically polarized wave.

In general, the polarization of a wave does not change over short distances. Therefore, transmitting and receiving antennas are oriented alike, especially if they are separated by short distances.

Over long distances, polarization changes. The change is usually small at low frequencies, but quite drastic at high frequencies. (For radar transmissions, a received signal is actually a wave reflected from an object. Since signal polarization varies with the type of object, no set position of the receiving antenna is correct for all returning signals). Where separate antennas are used for transmitting and receiving, the receiving antenna is generally polarized in the same direction as the transmitting antenna.

When the transmitting antenna is close to the ground, it should be polarized vertically, because vertically polarized waves produce a greater signal strength along the earth's surface. On the other hand, when the transmitting antenna is high above the ground, it should be horizontally polarized to get the greatest signal strength possible to the earth's surface.

RADIATION OF ELECTROMAGNETIC ENERGY

Various factors in the antenna circuit affect the radiation of electromagnetic energy. In figure 2-3, for example, if an alternating current is applied to the A end of wire antenna AB, the wave will travel along the wire until it reaches the B end. Since the B end is free, an open circuit exists and the wave cannot travel further. This is a point of high impedance. The wave bounces back (reflects) from this point of high impedance and travels toward the starting point, where it is again reflected. Theoretically, the energy of the wave should be gradually dissipated by the resistance of the wire during this back-and-forth motion (oscillation). However, each time the wave reaches the starting point, it is reinforced by an impulse of energy sufficient to replace the energy lost during its travel along the wire. This results in continuous oscillations of energy along the wire and a high voltage at the A end of the wire. These oscillations move along the antenna at a rate equal to the frequency of the rf voltage and are sustained by properly timed impulses at point A.

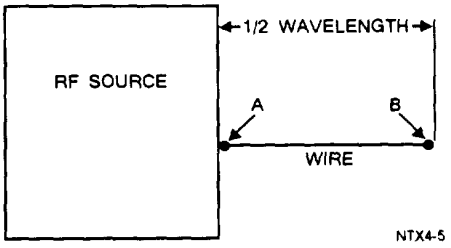


Figure 2-3.—Antenna and rf source.

The rate at which the wave travels along the wire is constant at approximately 300,000,000 meters per second. The length of the antenna must be such that a wave will travel from one end to the other and back again during the period of 1 cycle of the rf voltage. The distance the wave travels during the period of 1 cycle is known as the wavelength. It is found by dividing the rate of travel by the frequency.

Look at the current and voltage distribution on the antenna in figure 2-4. A maximum movement of electrons is in the center of the antenna at all times; therefore, the center of the antenna is at a low impedance.

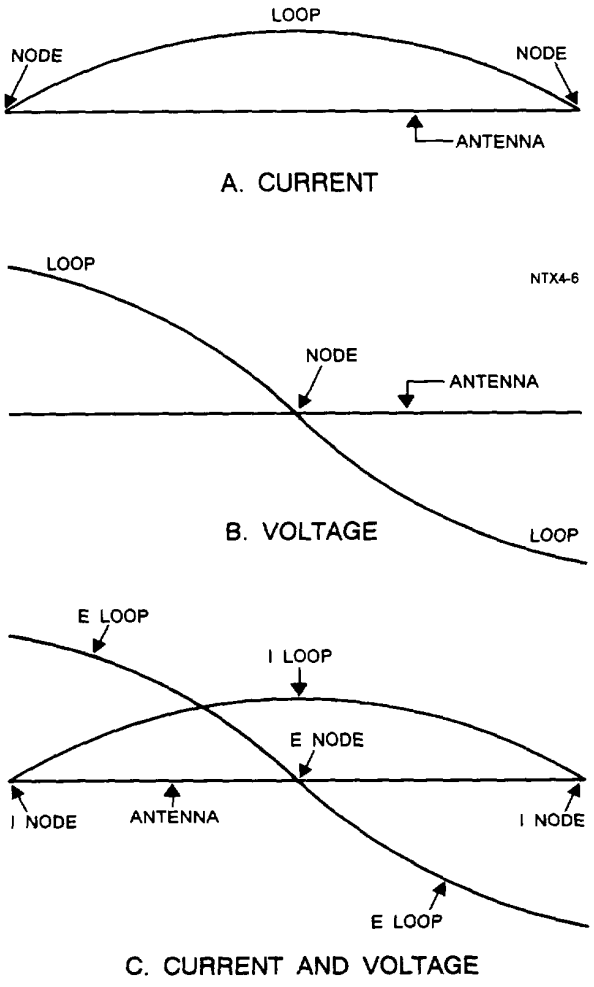


Figure 2-4.—Standing waves of current and voltage on an antenna.

This condition is called a STANDING WAVE of current. The points of high current and high voltage are known as current and voltage LOOPS. The points of minimum current and minimum voltage are known as current and voltage NODES. View A shows a current loop and two current nodes. View B shows two voltage loops and a voltage node. View C shows

the resultant voltage and current loops and nodes. The presence of standing waves describes the condition of resonance in an antenna. At resonance, the waves travel back and forth in the antenna, reinforcing each other, and are transmitted into space at maximum radiation. When the antenna is not at resonance, the waves tend to cancel each other and energy is lost in the form of heat.

RADIATION TYPES AND PATTERNS

A logical assumption is that energy leaving an antenna radiates equally over 360 degrees. This is not the case for every antenna.

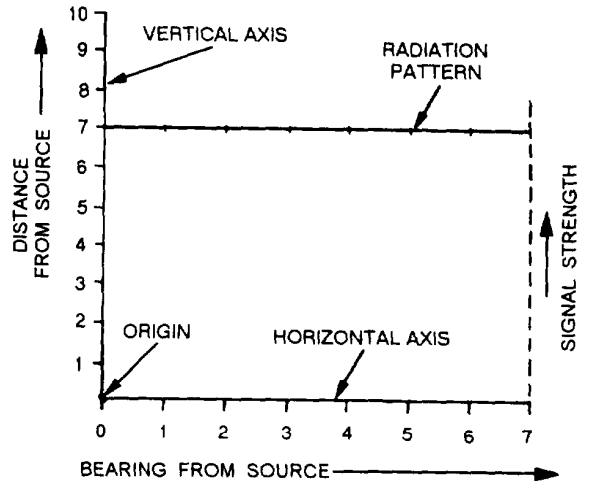
The energy radiated from an antenna forms a field having a definite RADIATION PATTERN. The radiation pattern for any given antenna is determined by measuring the radiated energy at various angles at constant distances from the antenna and then plotting the energy values on a graph. The shape of this pattern depends on the type of antenna being used.

Some antennas radiate energy equally in all directions. Radiation of this type is known as ISOTROPIC RADIATION. The sun is a good example of an isotropic radiator. If you were to measure the amount of radiated energy around the sun's circumference, the readings would all be fairly equal (fig. 2-5).

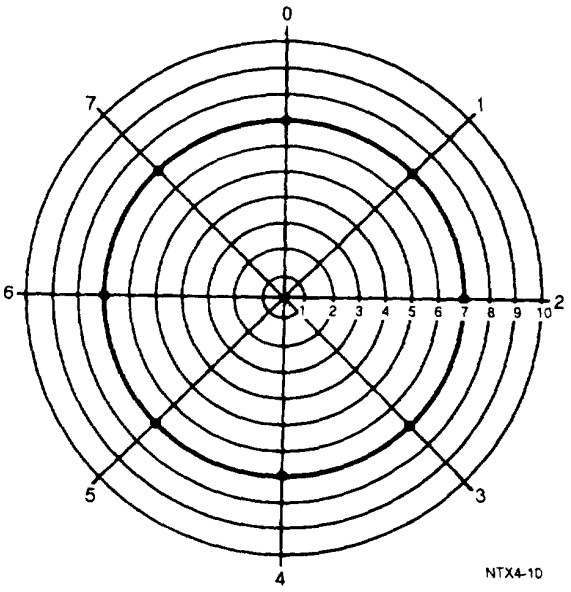
Most radiators emit (radiate) energy more strongly in one direction than in another. These radiators are referred to as ANISOTROPIC radiators. A flashlight is a good example of an anisotropic radiator (fig. 2-6). The beam of the flashlight lights only a portion of the space surrounding it. The area behind the flashlight remains unlit, while the area in front and to either side is illuminated.

MAJOR AND MINOR LOBES

The pattern shown in figure 2-7, view B, has radiation concentrated in two lobes. The radiation intensity in one lobe is considerably stronger than in the other. The lobe toward point X is called a MAJOR LOBE; the other is a MINOR LOBE. Since the complex radiation patterns associated with antennas frequently contain several lobes of varying intensity,



A. RECTANGULAR-COORDINATE GRAPH



B. POLAR-COORDINATE GRAPH

Figure 2-5.—Isotropic radiation graphs.

you should learn to use the appropriate terminology, In general, major lobes are those in which the greatest amount of radiation occurs. Minor lobes are those in which the least amount of radiation occurs.

ANTENNA LOADING

There will be times when you may want to use one antenna system to transmit on several different frequencies. Since the antenna must always be in resonance with the applied frequency, you must either lengthen it or shorten it to produce the required

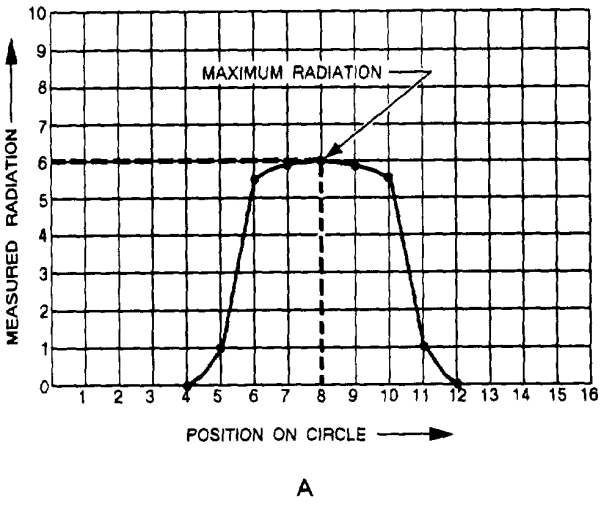


Figure 2-6.—Anisotropic radiator.

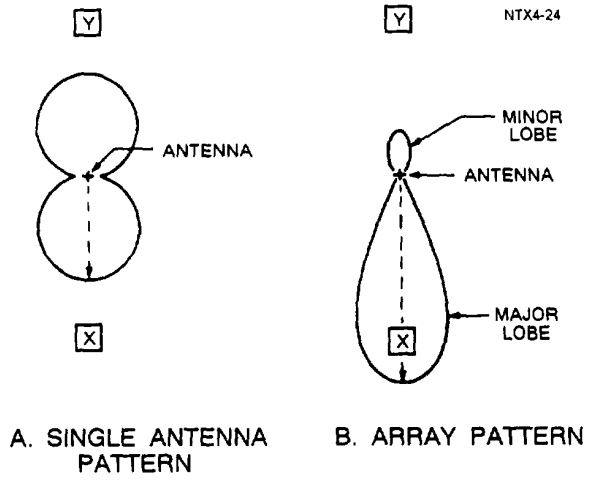


Figure 2-7.—Major and minor lobes.

resonance. Changing the antenna dimensions physically is impractical, but changing them electrically is relatively simple. To change the electrical length of an antenna, you can insert either an inductor or a capacitor in series with the antenna. This is shown in figure 2-8, views A and B. Changing the electrical length by this method is known as LUMPED-IMPEDANCE TUNING or LOADING. If the antenna is too short for the wavelength being used, it will be resonant at a higher frequency. Therefore, it offers a capacitive reactance at the excitation frequency. This capacitive reactance can be compensated for by introducing a lumped inductive reactance, as shown in view A. Similarly, if the antenna is too long for the transmitting frequency, it will be resonant at a lower frequency and offers an inductive reactance. Inductive reactance can be compensated for by introducing a lumped capacitive reactance, as shown in view B. An antenna with normal loading is represented in view C.

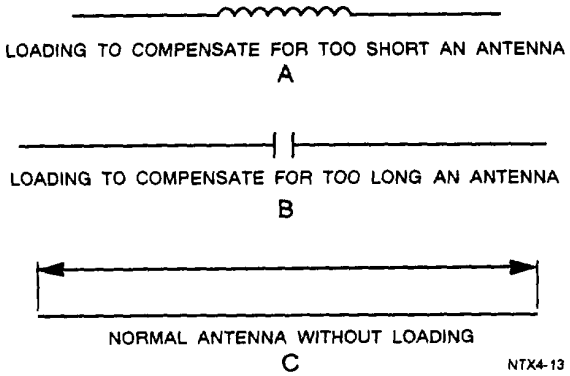


Figure 2-8.—Electrical antenna loading.

GROUND EFFECTS

As we discussed earlier, ground losses affect radiation patterns and cause high signal losses for some frequencies. Such losses can be greatly reduced if a good conducting ground is provided in the vicinity of the antenna. This is the purpose of the GROUND SCREEN (fig. 2-9, view A) and COUNTERPOISE (fig. 2-9, view B).

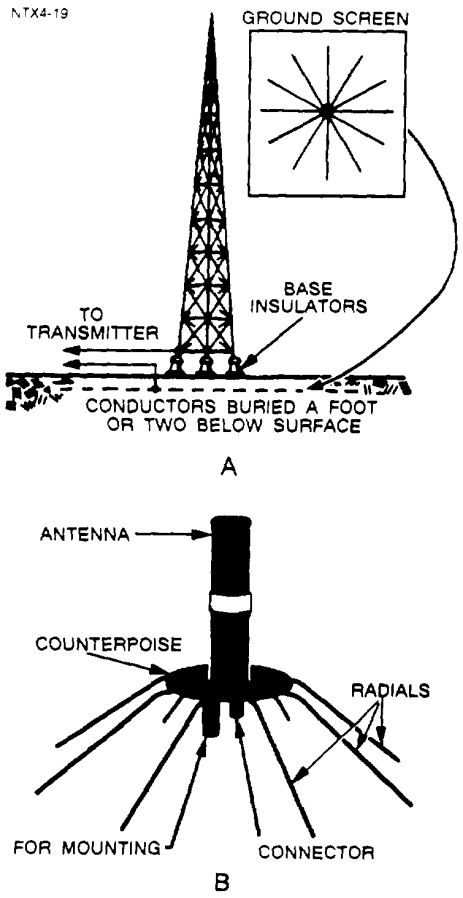


Figure 2-9.—Ground screen and counterpoise.

The ground screen in view A is composed of a series of conductors arranged in a radial pattern and buried 1 or 2 feet below the surface of the earth. These conductors, each usually 1/2 wavelength long, reduce ground absorption losses in the vicinity of the antenna.

A counterpoise (view B) is used when easy access to the base of the antenna is necessary. It is also used when the area below the antenna is not a good conducting surface, such as solid rock or ground that is sandy. The counterpoise serves the same purpose as the ground screen but is usually elevated above the earth. No specific dimensions are necessary for a counterpoise, nor is the number of wires particularly critical. The primary requirement is that the counterpoise be insulated from ground and form a grid of reflector elements for the antenna system.

COMMUNICATIONS ANTENNAS

Some antennas can be used in both shore-based and ship-based applications. Others, however, are designed to be used primarily in one application or the other. The following paragraphs discuss, by frequency range, antennas used for shore-based communications.

VERY LOW FREQUENCY (VLF)

The main difficulty in vlf and lf antenna design is the physical disparity between the maximum practical size of the antenna and the wavelength of the frequency it must propagate. These antennas must be large to compensate for wavelength and power handling requirements (0.25 to 2 MW), Transmitting antennas for vlf have multiple towers 600 to 1500 feet high, an extensive flat top for capacitive loading, and a copper ground system for reducing ground losses. Capacitive top-loading increases the bandwidth characteristics, while the ground plane improves radiation efficiency.

Representative antenna configurations are shown in figures 2-10 through 2-12. Variations of these basic antennas are used at the majority of the Navy vlf sites.

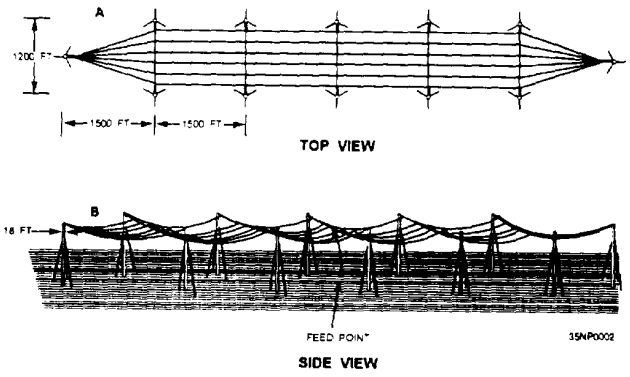


Figure 2-10.—Triatic-type antenna.

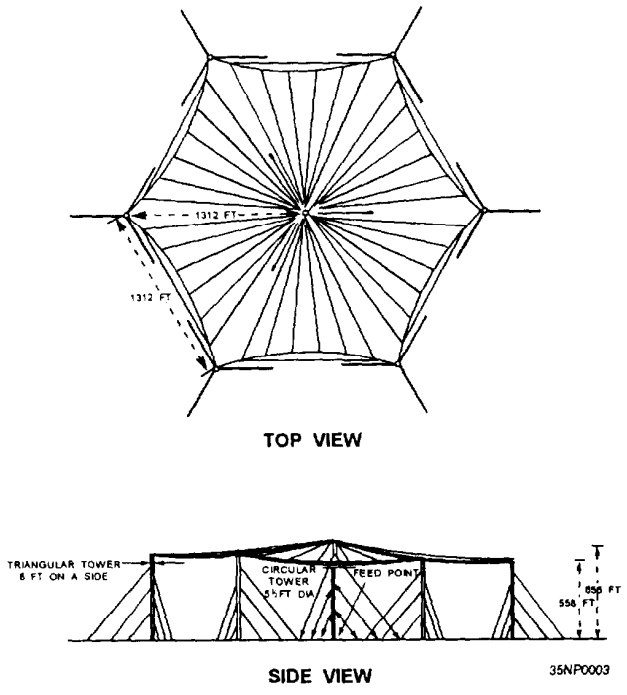


Figure 2-11.—Goliath-type antenna.

LOW FREQUENCY (LF)

Antennas for lf are not quite as large as antennas for vlf, but they still occupy a large surface area. Two examples of If antenna design are shown in figures 2-13 and 2-14. The Pan polar antenna (fig. 2-1 3) is an umbrella top-loaded monopole. It has three loading loops spaced 120 degrees apart, interconnected between the tower guy cables. Two of the loops terminate at ground, while the other is used as a feed. The NORD antenna (fig. 2-14), based on the the folded-unipole principle, is a vertical tower radiator grounded at the base and fed by one or more wires connected to the top of the tower. The three top loading wires extend from the top of the antenna at 120-degree intervals to three terminating towers. Each loading wire has a length approximately equal to the height of the main tower plus 100 feet. The top loading wires are insulated from ground and their tower supports are one-third the height of the transmitting antenna.

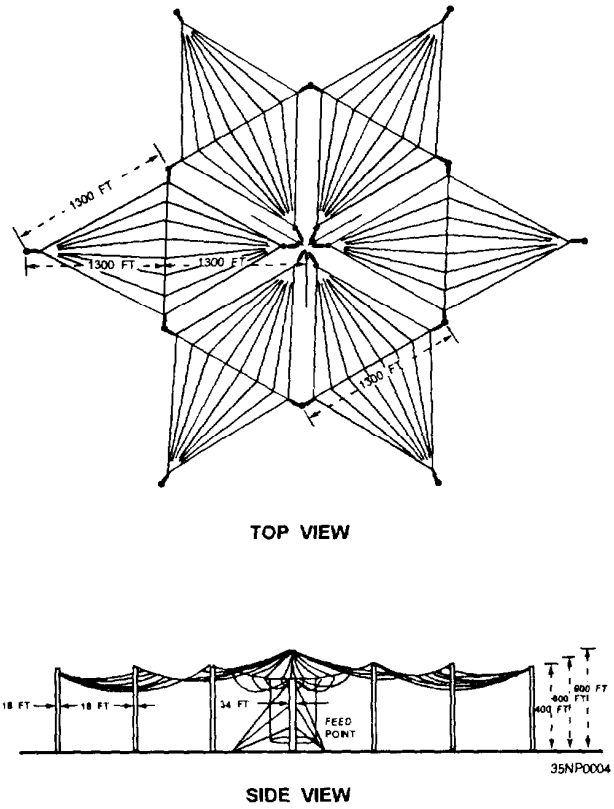


Figure 2-12.—Trideco-type antenna.

HIGH FREQUENCY (HF)

High-frequency (hf) radio antenna systems are used to support many different types of circuits, including ship-to-shore, point-to-point, and ground-to-air broadcast. These diverse applications require the use of various numbers and types of antennas that we will review on the following pages.

Yagi

The Yagi antenna is an end-fired parasitic array. It is constructed of parallel and coplaner dipole elements arranged along a line perpendicular to the axis of the dipoles, as illustrated in figure 2-15. The most limiting characteristic of the Yagi antenna is its extremely narrow bandwidth. Three percent of the center frequency is considered to be an acceptable bandwidth ratio for a Yagi antenna. The width of the array is determined by the lengths of the elements. The length of each element is approximately one-half

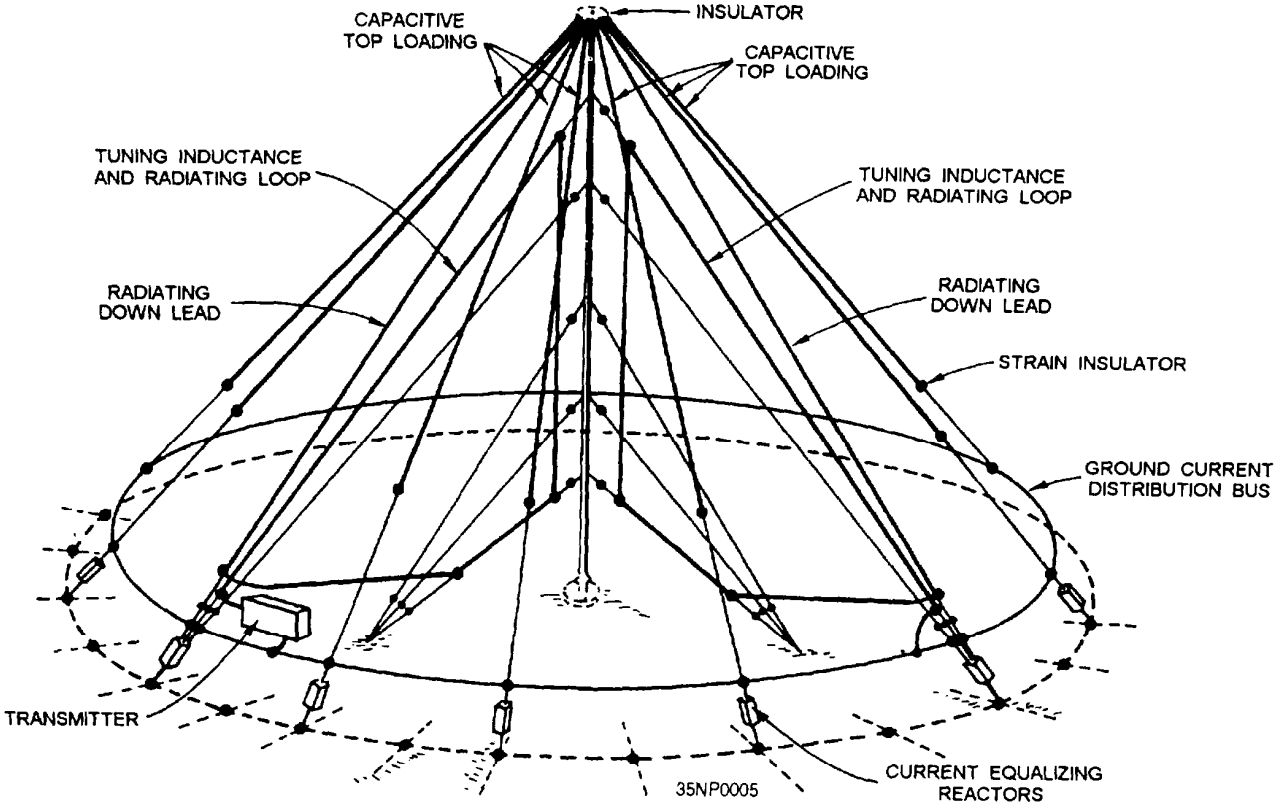


Figure 2-13.—Pan polar antenna.

wavelength, depending on its intended use (driver, reflector, or director). The required length of the array depends on the desired gain and directivity. Typically, the length will vary from 0.3 wavelength for three-element arrays, to 3 wavelengths for arrays with numerous elements. For hf applications, the maximum practical array length is 2 wavelengths. The array's height above ground will determine its vertical radiation angle. Normally, array heights vary from 0.25 to 2.5 wavelengths. The dipole elements are usually constructed from tubing, which provides for better gain and bandwidth characteristics and provides sufficient mechanical rigidity for self-support. Yagi arrays of four elements or less are not structurally complicated. Longer arrays and arrays for lower frequencies, where the width of the array exceeds 40 feet, require elaborate booms and supporting structures. Yagi arrays may be either fixed-position or rotatable.

LOG-PERIODIC ANTENNAS (LPAs)

An antenna arranged so the electrical length and spacing between successive elements causes the input impedance and pattern characteristics to be repeated periodically with the logarithm of the driving frequency is called a LOG-PERIODIC ANTENNA (LPA). The LPA, in general, is a medium-power, high-gain, moderately-directive antenna of extremely broad bandwidth. Bandwidths of up to 15:1 are possible, with up to 15 dB power gain. LPAs are rather complex antenna systems and are relatively expensive. The installation of LPAs is normally more difficult than for other hf antennas because of the tower heights involved and the complexity of suspending the radiating elements and feedlines from the towers.

Vertical Monopole LPA

The log-periodic vertical monopole antenna (fig. 2-16) has the plane containing the radiating elements in a vertical field. The longest element is approximately one-quarter wavelength at the lower cutoff frequency. The ground system for the monopole arrangement provides the image equivalent of the other quarter wavelength for the half-dipole radiating elements. A typical vertical monopole designed to

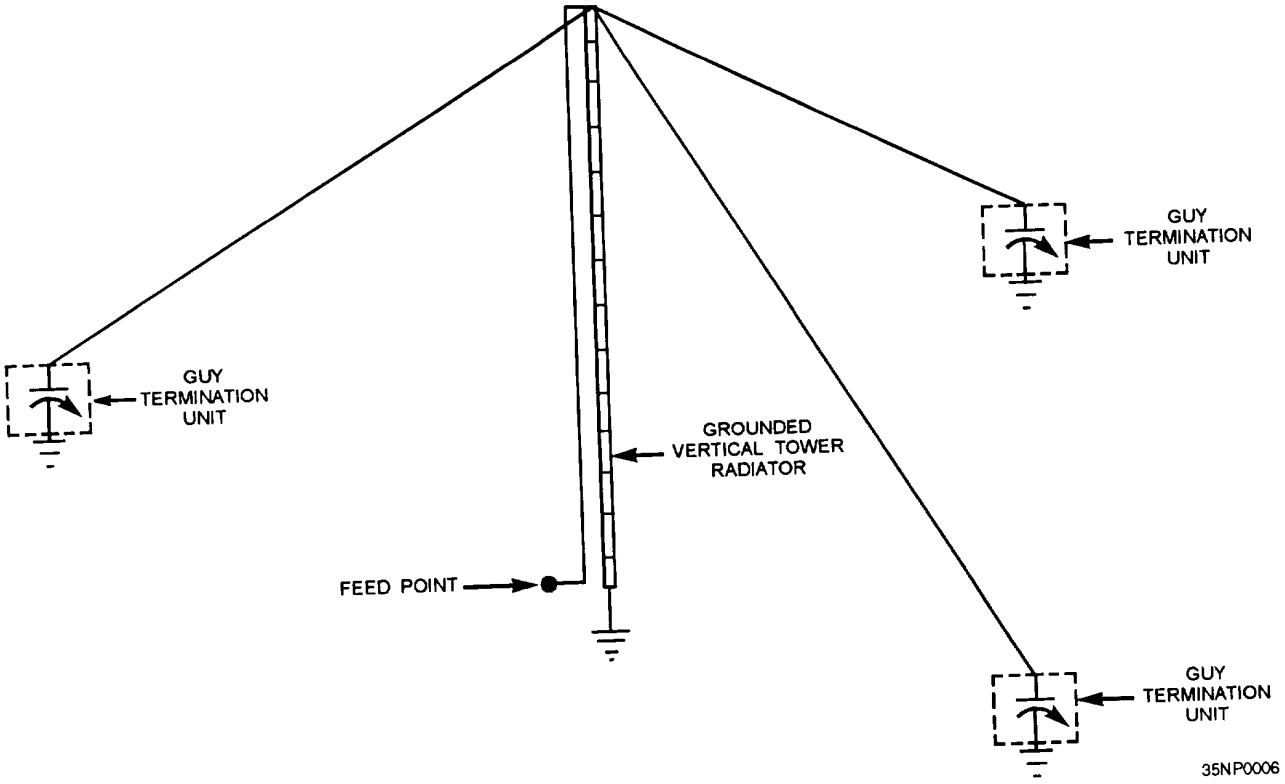


Figure 2-14.—NORD antenna.

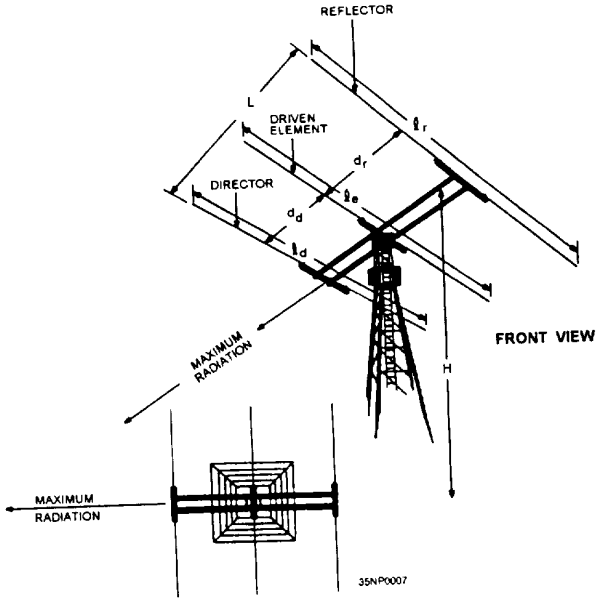


Figure 2-15.—Yagi antenna.

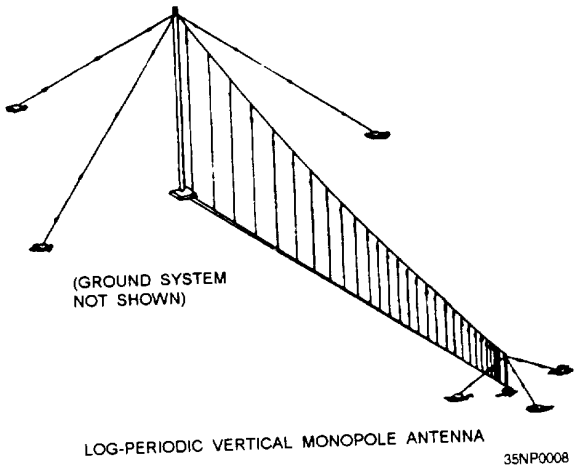


Figure 2-16.—Log-periodic vertical monopole antenna.

cover a frequency range of 2 to 30 MHz requires one tower approximately 140 feet high and an antenna length of around 500 feet, with a ground system that covers approximately 3 acres of land in the immediate vicinity of the antenna.

Sector Log-Periodic Array

This version of a vertically polarized fixed-azimuth LPA consists of four separate curtains supported by a common central tower, as shown in figure 2-17. Each of the four curtains operates independently, providing antennas for a minimum of four transmit or receive systems. and a choice of sector coverage. The four curtains are also capable of radiating a rosette pattern of overlapping sectors for full coverage, as shown by the radiation pattern in figure 2-17. The central supporting tower is constructed of steel and may range to approximately 250 feet in height, with the length of each curtain reaching 250 feet, depending on its designed operating frequencies. A sector antenna that uses a ground plane designed to cover the entire hf spectrum takes up 4 to 6 acres of land area.

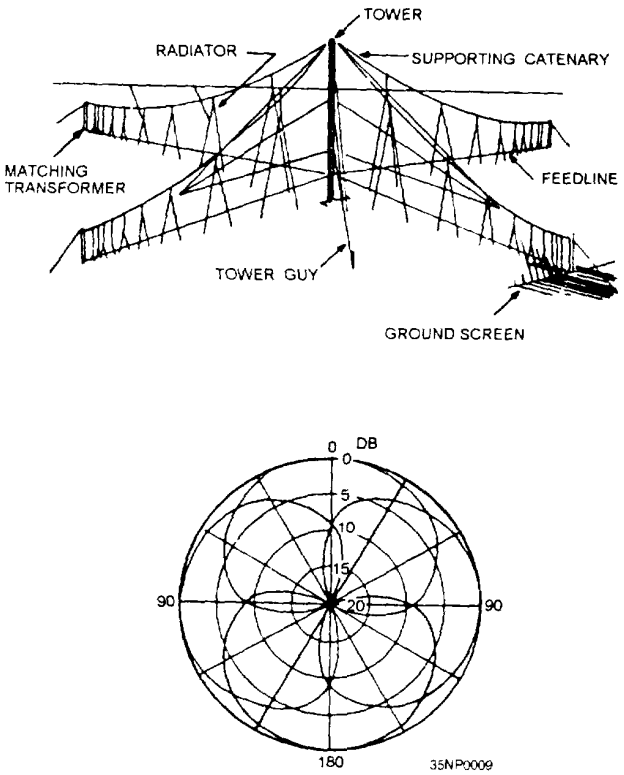


Figure 2-17.—Sector LPA and its horizontal radiation pattern.

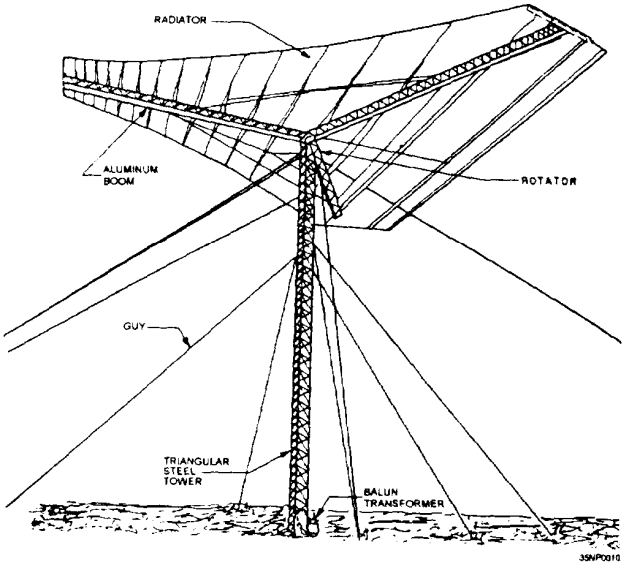


Figure 2-18.—Rotatable log-periodic antenna.

Rotatable LPA (RLPA)

RLPAs (fig. 2-18) are commonly used in ship-to-shore-to-ship and in point-to-point ecm-u-nunications. Their distinct advantage is their ability to rotate 360 degrees. RLPAs are usually constructed with either tubular or wire antenna elements. The RLPA in figure 2-18 has wire elements strung on three aluminum booms of equal length, spaced equally and arranged radially about a central rotator on top of a steel tower approximately 100 feet high. The frequency range of this antema is 6 to 32 MHz. The gain is 12 dB with respect to isotropic antennas. Power handling capability is 20 kw average, and vswr is 2:1 over the frequency range.

INVERTED CONE ANTENNA

Inverted cone antennas are vertically polarized, omnidirectional, and have an extremely broad bandwidth. They are widely used for ship-to-shore and ground-to-air communications. Inverted cone antennas are installed over a radial ground plane system and are supported by poles, as shown in figure 2-19. The equally-spaced vertical radiator wires terminate in a feed ring assembly located at the bottom center, where a 50-ohm coaxial transmission line feeds the antenna. Inverted cones usually have gains of 1 to 5 dB above isotropic antennas, with a vswr not

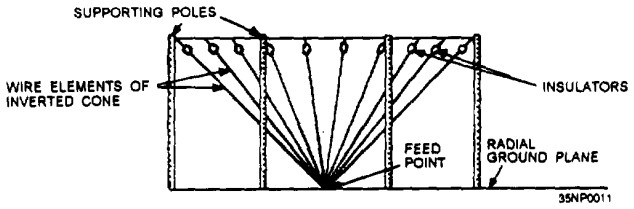


Figure 2-19.—Inverted cone antenna.

greater than 2:1. They are considered medium- to high-power radiators, with power handling capabilities of 40 kW average power.

CONICAL MONOPOLE ANTENNA

Conical monopoles are used extensively in hf communications. A conical monopole is an efficient broadband, vertically polarized, omnidirectional antenna in a compact size. Conical monopoles are shaped like two truncated cones connected base-to-base. The basic conical monopole configuration, shown in figure 2-20, is composed of equally-spaced wire radiating elements arranged in a circle around an aluminum center tower. Usually, the radiating elements are connected to the top and bottom discs, but on some versions, there is a center waist disc where the top and bottom radiators are connected. The conical monopole can handle up to 40 kW of average power. Typical gain is -2 to +2 dB, with a vswr of up to 2.5:1.

RHOMBIC ANTENNA

Rhombic antennas can be characterized as high-power, low-angle, high-gain, horizontally-polarized, highly-directive, broadband antennas of simple, inexpensive construction. The rhombic antenna (fig. 2-21) is a system of long-wire radiators that depends on radiated wave interaction for its gain and directivity. A properly designed rhombic antenna presents to the transmission line an input impedance insensitive to frequency variations up to 5:1. It maintains a power gain above 9 dB anywhere within a 2:1 frequency variation. At the design-center frequency, a gain of 17 dB is typical. The radiation pattern produced by the four radiating legs of a rhombic antenna is modified by reflections from the earth under, and immediately in front of, the antenna. Because of the importance of these ground

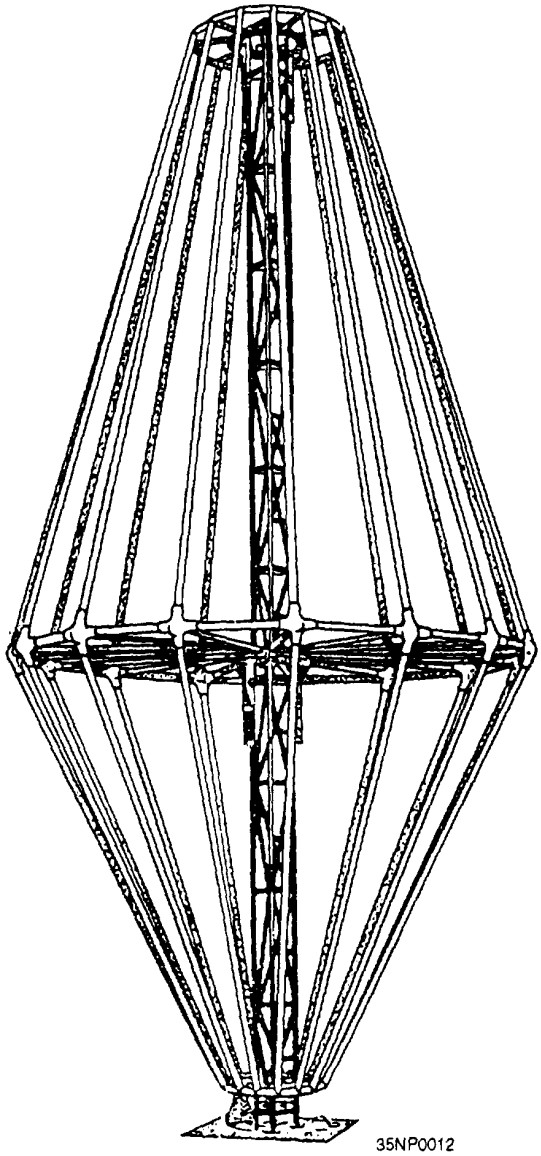


Figure 2-20.—Conical monopole antenna.

reflections in the proper formation of the main lobe, the rhombic should be installed over reasonably smooth and level ground. The main disadvantage of the rhombic antenna is the requirement for a large land area, usually 5 to 15 acres.

QUADRANT ANTENNA

The hf quadrant antenna (fig. 2-22) is a special-purpose receiving antenna used in ground-to-air-to-ground communications. It is unique among horizontally-polarized antennas because its

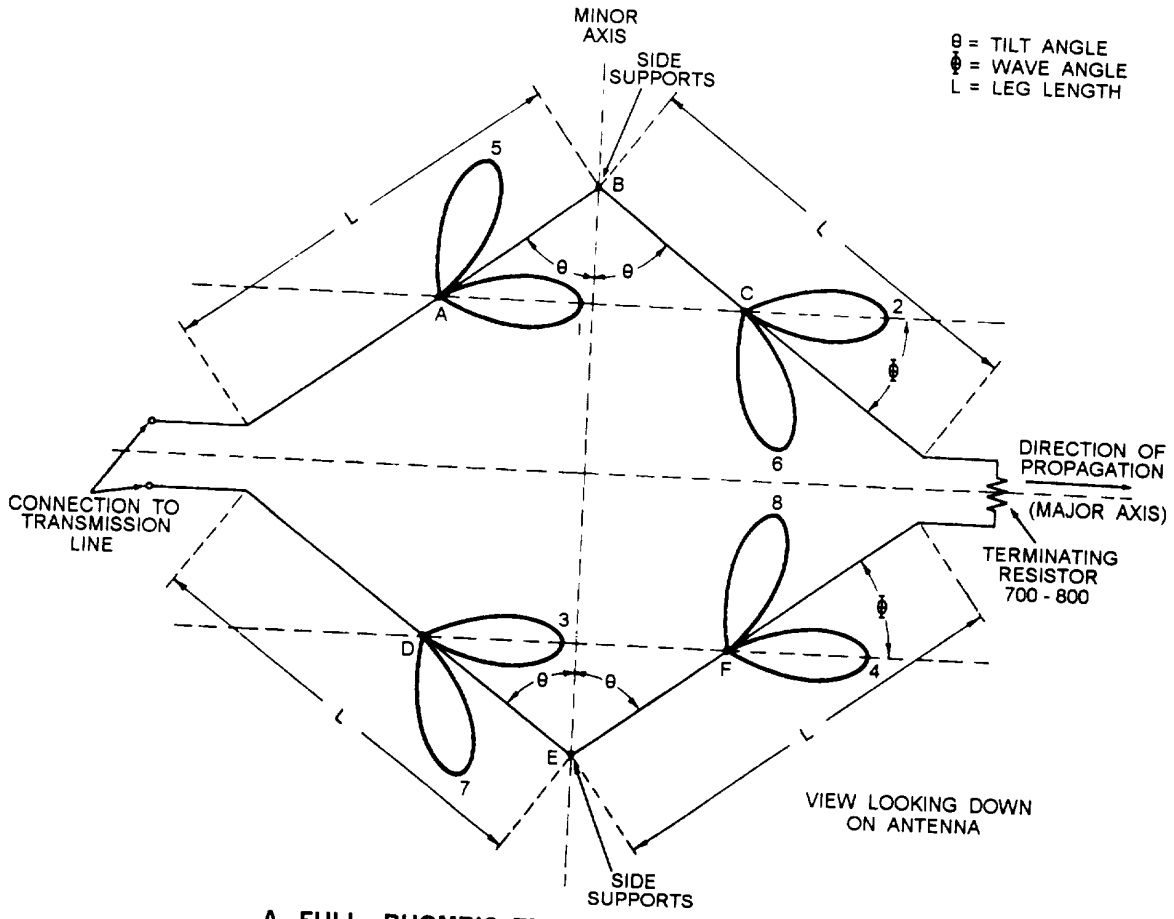


Figure 2-21.—Three-wire rhombic antenna.

element arrangement makes possible a radiation pattern resembling that of a vertically-polarized, omnidirectional antenna. Construction and installation of this antenna is complex because of the physical relationships between the individual elements and the requirement for a separate transmission line for each dipole. Approximately 2.2 acres of land are required to accommodate the quadrant antenna.

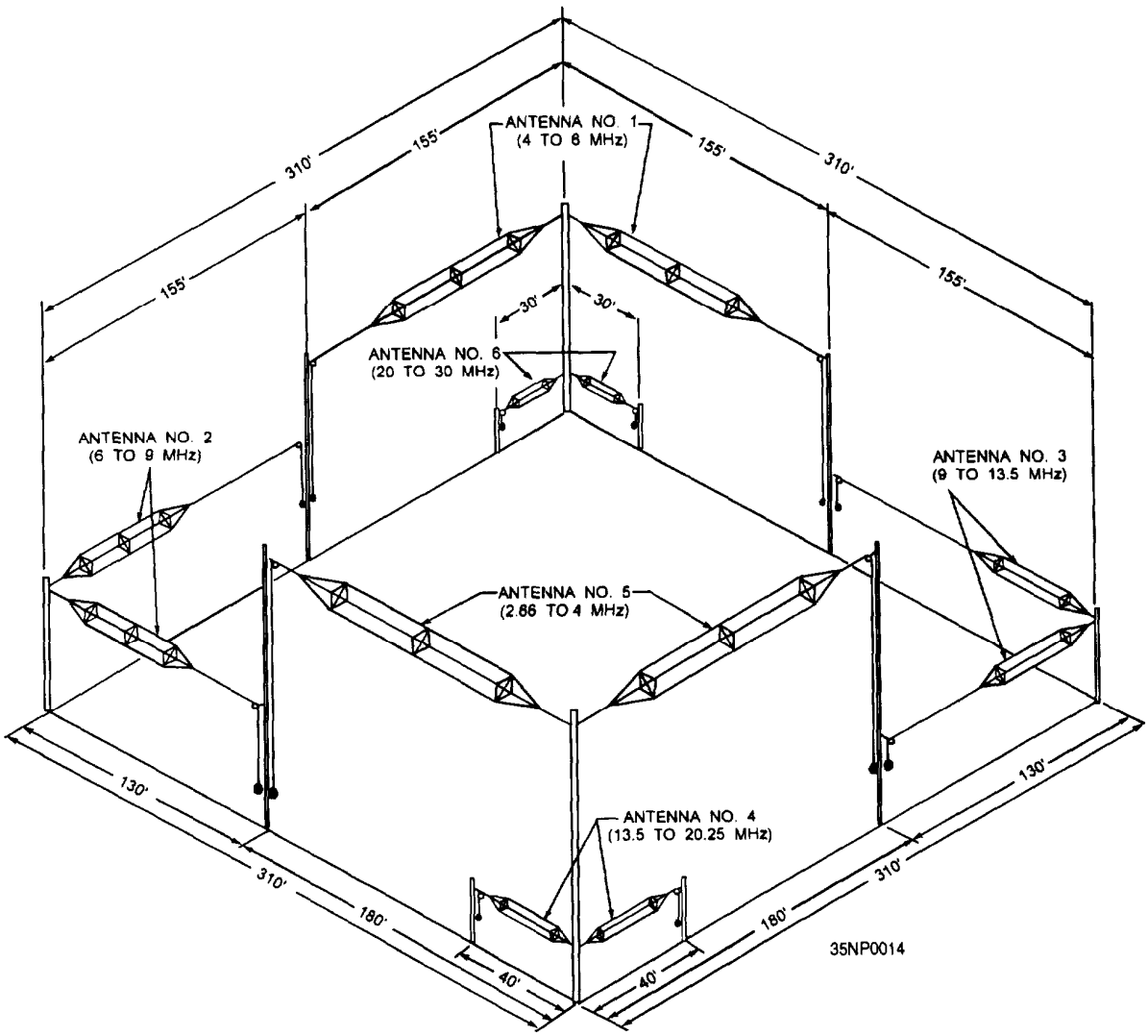


Figure 2-22.—Quadrant antenna.

WHIP ANTENNAS

Hf whip antennas (fig. 2-23) are vertically-polarized omnidirectional monopoles that are used for short-range, ship-to-shore and transportable communications systems. Whip antennas are made of tubular metal or fiberglass, and vary in length from 12 feet to 35 feet, with the latter being the most prevalent. Although whips are not considered as highly efficient antennas, their ease of installation and low cost provide a compromise for receiving and low-to-medium power transmitting installations.

The self-supporting feature of the whip makes it particularly useful where space is limited. Whips can be tilted, a design feature that makes them suited for use along the edges of aircraft carrier flight decks. Aboard submarines, they can be retracted into the sail structure.

Most whip antennas require some sort of tuning system and a ground plane to improve their radiation efficiency throughout the hf spectrum. Without an antenna tuning system, whips generally have a narrow bandwidth and are limited in their power handling

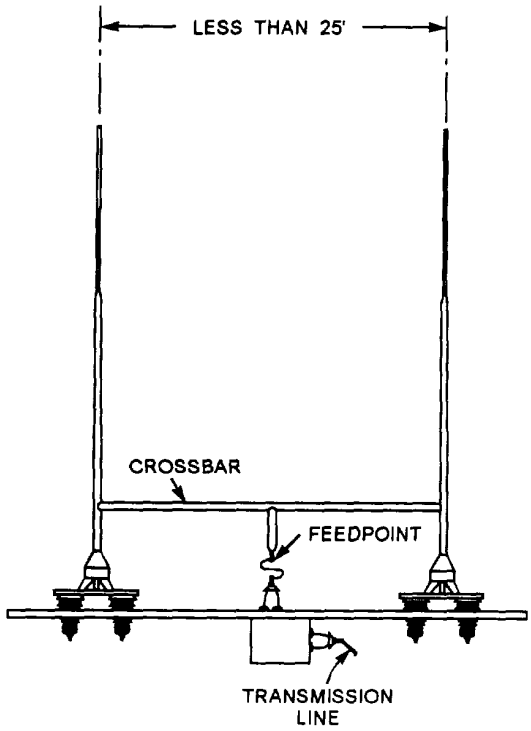


Figure 2-23.—Whip antennas.

capabilities. Power ratings for most whips range from 1 to 5 kW PEP.

WIRE-ROPE FAN ANTENNAS

Figure 2-24 shows a five-wire vertical fan antenna. This is a broadband antenna composed of five wires,

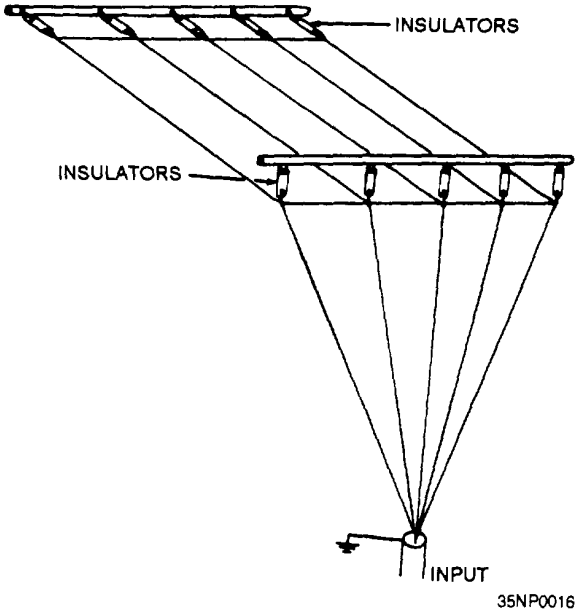


Figure 2-24.—Vertical fan antenna.

each cut for one-quarter wavelength at the lowest frequency to be used. The wires are fanned 30 degrees between adjacent wires. The fan antenna provides satisfactory performance and is designed for use as a random shipboard antenna in the hf range (2-30 MHz).

DISCAGE ANTENNA

The discage antenna (fig. 2-25) is a broadband omnidirectional antenna. The diseage structure consists of two truncated wire rope cones attached base-to-base and supported by a central mast. The lower portion of the structure operates as a cage monopole for the 4- to 12-MHz frequency range. The upper portion operates as a discone radiator in the 10- to 30-MHz frequency range. Matching networks limit the vswr to not greater than 3:1 at each feed point. Vinyl-covered phosphor bronze wire rope is used for the wire portions. The support mast and other portions are aluminum.

VHF/UHF

At vhf and uhf frequencies, the shorter wavelength makes the physical size of the antenna relatively small. Aboard ship these antennas are installed as high as

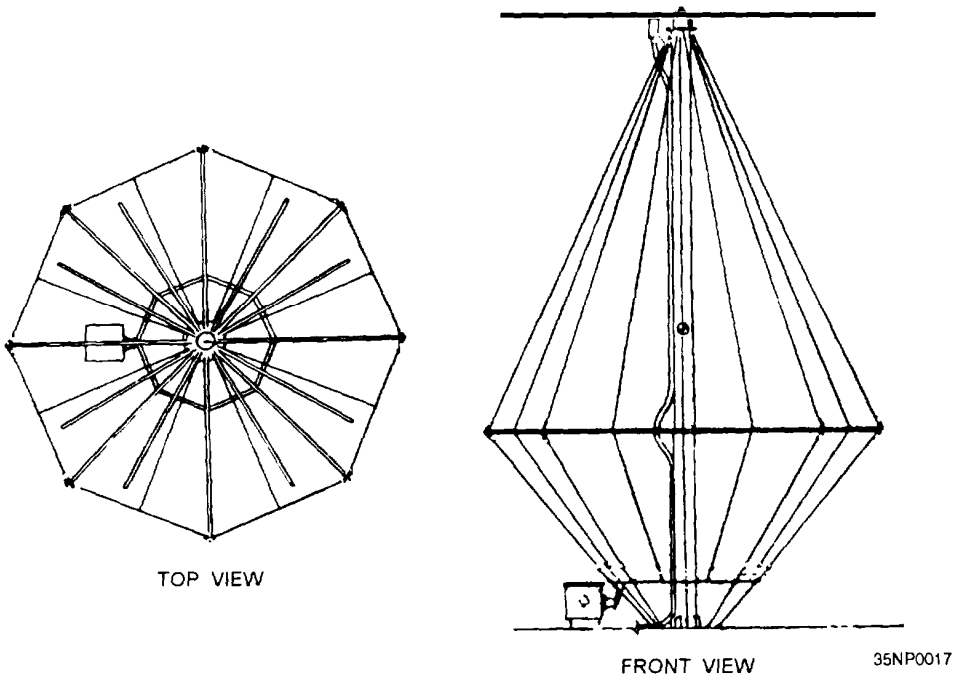


Figure 2-25.—AS-2802/SCR discage antenna.

possible and away from any obstructions. The reason for the high installation is that vertical conductors, such as masts, rigging, and cables in the vicinity, cause unwanted directivity in the radiation pattern.

For best results in the vhf and uhf ranges, both transmitting and receiving antennas must have the same polarization. Vertically polarized antennas (primarily dipoles) are used for all ship-to-ship, ship-to-shore, and air-to-ground vhf and uhf communications.

The following paragraphs describe the most common uhf/vhf dipole antennas. All the examples are vertically-polarized, omnidirectional, broadband antennas.

Biconical Dipole

The biconical dipole antenna (fig. 2-26) is designed for use at a normal rf power rating of around 250 watts, with a vswr not greater than 2:1. All major components of the radiating and support structures are aluminum. The central feed section is protected and waterproofed by a laminated fiberglass cover.

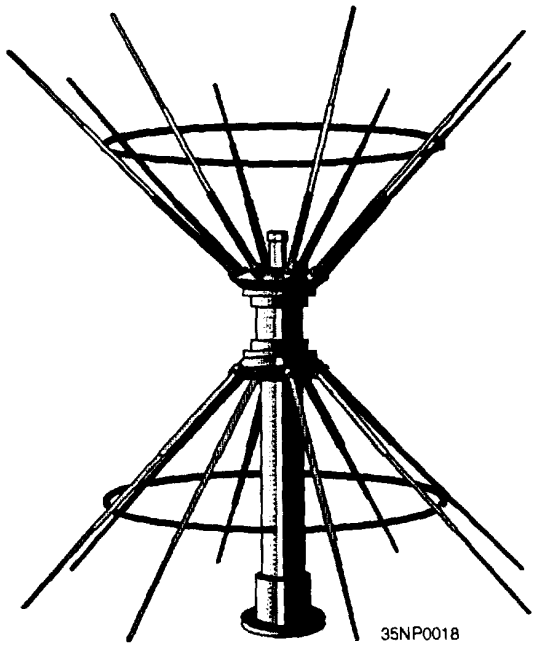


Figure 2-26.—AS-2811/SCR biconical dipole antenna.

Center-Fed Dipole

The center-fed dipole (fig. 2-27) is designed for use at an average power rating of 100 watts. All major components of the radiating and support structures are aluminum. The central feed section and radiating elements are protected by a laminated fiberglass cover. Center-fed dipole antennas range from 29 to 47 inches in height and have a radiator diameter of up to 3 inches.

Coaxial Dipole

Figure 2-28 shows two types of coaxial dipoles. The coaxial dipole antenna is designed for use in the uhf range, with an rf power rating of 200 watts. The AT-150/SRC (fig. 2-28, view A) has vertical radiating elements and a balun arrangement that electrically balances the antenna to ground.

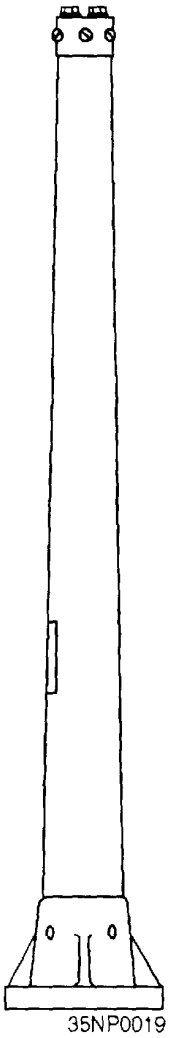


Figure 2-27.—AS-2809/RC center-fed dipole antenna.

Figure 2-28, view B, shows an AS-390/SRC antenna assembly. This antenna is an unbalanced broadband coaxial stub antenna. It consists of a radiator and a ground plane. The ground plane (or counterpoise) consists of eight elements bent downward 37 degrees from horizontal. The lower ends of the elements form points of a circle 23 inches in diameter. The lower section of the radiator assembly contains a stub for adjusting the input impedance of the antenna. The antenna is vertically polarized, with an rf power rating of 200 watts, and a vswr not greater than 2:1.

SATELLITE SYSTEMS

The Navy Satellite Communication System (SATCOM) provides communications links, via satellites, between designated mobile units and shore sites. These links supply worldwide communications coverage. The following paragraphs describe some of the more common SATCOM antenna systems to which you will be exposed.

AS-2815/SRR-1

The AS-2815/SSR-1 fleet broadcast receiving antenna (fig. 2-29) has a fixed 360-degree horizontal pattern with a maximum gain of 4 dB at 90 degrees from the antenna's horizontal plane. The maximum loss in the antenna's vertical pattern sector is 2 dB. The vswr is less than 1.5:1, referenced to 50 ohms. This antenna should be positioned to protect it from interference and possible front end burnout from radar and uhf transmitters.

ANTENNA GROUPS OE-82B/WSC-1(V) AND OE-82C/WSC-1(V)

Designed primarily for shipboard installations, these antenna groups interface with the AN/WSC-3 transceiver. The complete installation consists of an antenna, bandpass amplifier-filter, switching unit, and antenna control (figs. 2-30 and 2-31), Depending on requirements, one or two antennas may be installed to provide a view of the satellite at all times. The antenna assembly is attached to a pedestal that permits

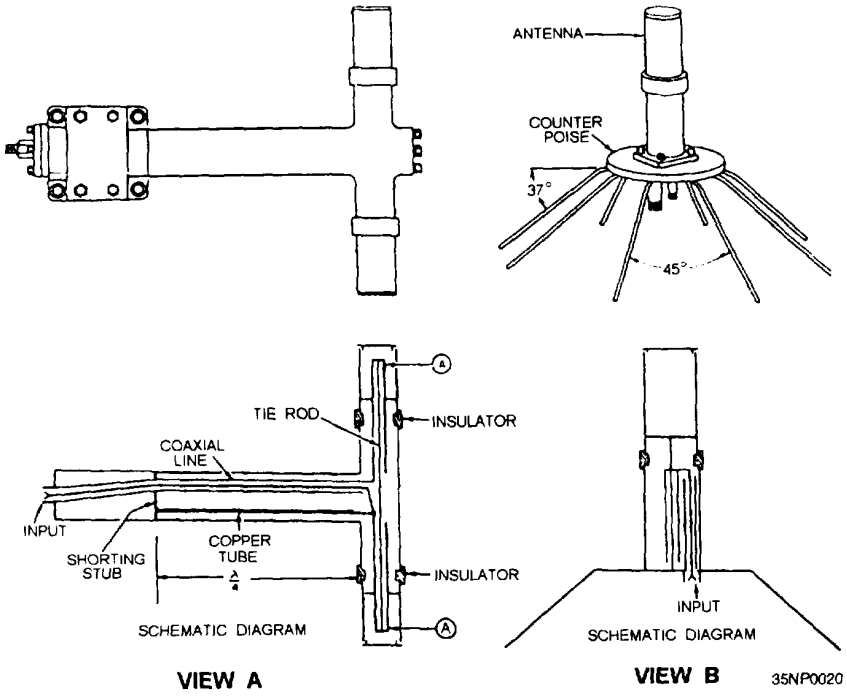


Figure 2-28.—Coaxial dipole.

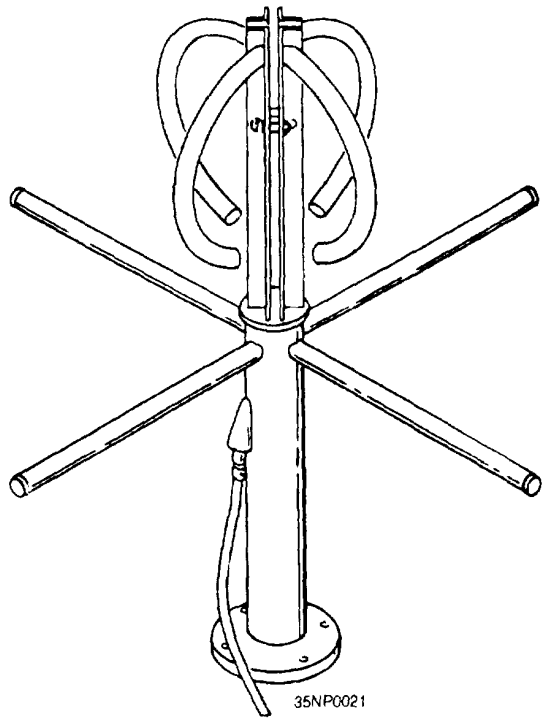


Figure 2-29.—AS-2815/SSR-1 fleet broadcast satellite receiving antenna.

it to rotate 360 degrees and to elevate from near horizontal to approximately 20 degrees beyond zenith (elevation angles from +2 to +110 degrees). The antenna tracks automatically in azimuth and manually in elevation. Frequency bands are 248-272 MHz for receive and 292-312 MHz for transmit. Polarization is right-hand circular for both transmit and receive. Antenna gain characteristics are nominally 12 dB in transmit and 11 dB in receive.

AN/WSC-5(V) SHORE STATION ANTENNA

The AN/WSC-5(V) shore station antenna (fig. 2-32) consists of four OE-82A/WSC-1(V) backplane assemblies installed on a pedestal. This antenna is intended for use with the AN/WSC-5(V) transceiver at major shore stations. The antenna is oriented manually and can be locked in position to receive maximum signal strength upon capture of the satellite signal. Hemispherical coverage is 0 to 110 degrees above the horizon. Polarization is right-hand circular in both transmit and receive. The antenna's operating frequency range is 240 to 318 MHz. With its mount,

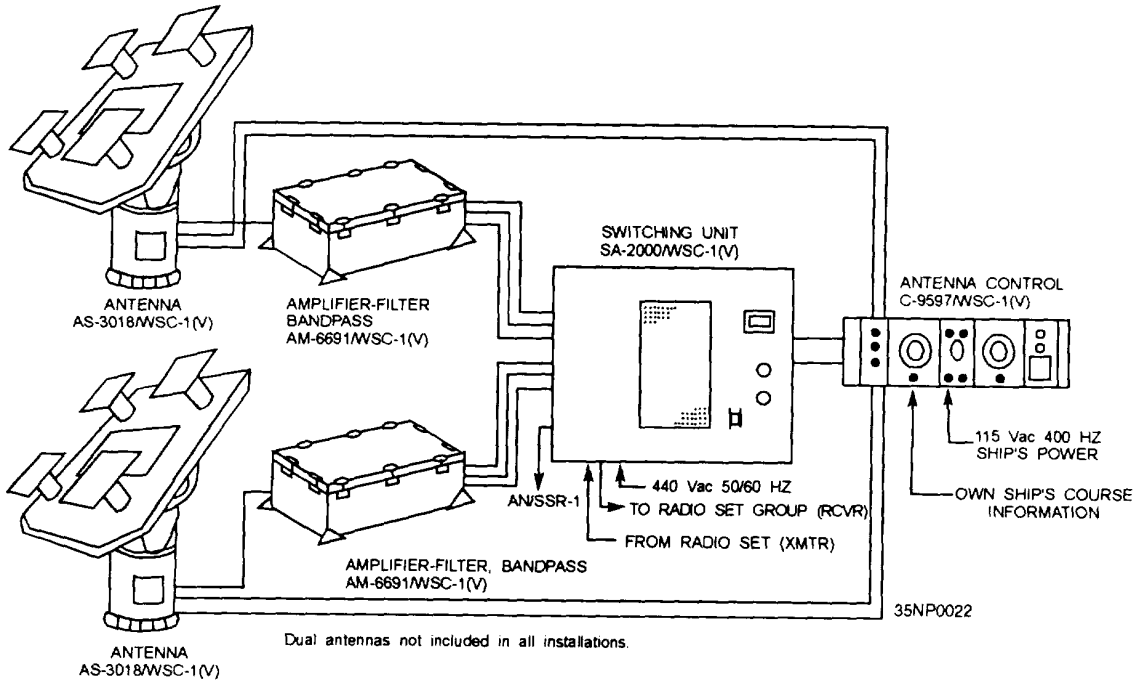


Figure 2-30.—OE-82/WSC-1(V) antenna group.

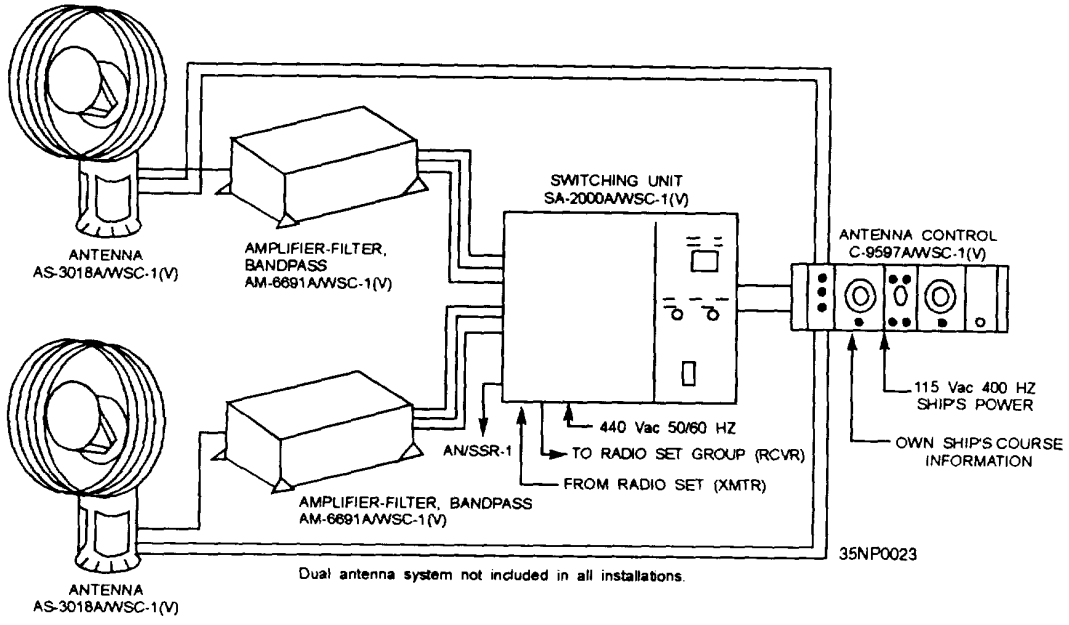


Figure 2-31.—OE-82C/WSC-1(V) antenna group.

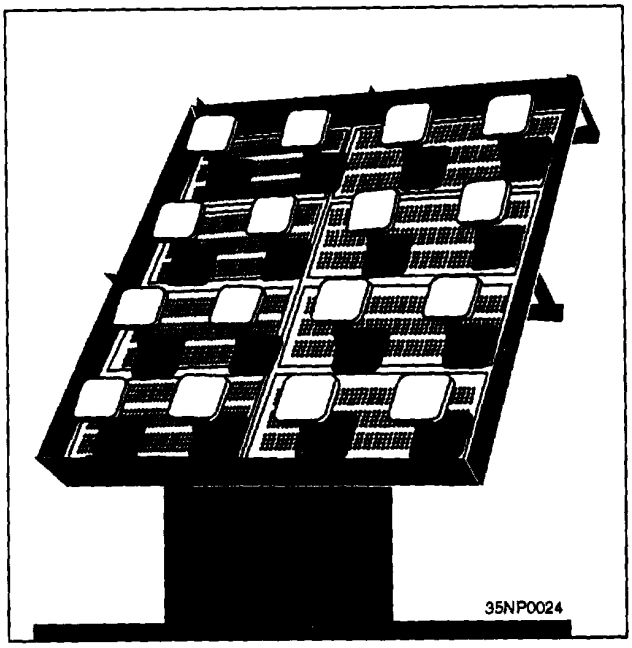


Figure 2-32.—OE-82A/WSC-1(V)/AN/WSC-5(V) shore station antenna.

the antenna weighs 2500 pounds and is 15 feet high, 10 feet wide, and 10 feet deep. The gain characteristics of this antenna are nominally 15 dB in transmit and 18 dB in receive.

ANDREW 58622 SHORE ANTENNA

The Andrew 58622 antenna (fig. 2-33) is a bifilar, 16-turn helical antenna right-hand circularly polarized, with gain varying between 11.2 and 13.2 dB in the 240-315 MKz frequency band. It has a 39-inch ground plate and is about 9 feet, 7 inches long. It can be adjusted manually in azimuth and elevation. This antenna is used at various shore installations, other than NCTAMS, for transmit and receive operations.

AN/WSC-6(V) SHF SATCOM ANTENNA

The antennas used on current shf SATCOM shipboard terminals are parabolic reflectors with casseegrain feeds. These antennas provide for LPI (low probability of intercept), with beamwidths less than 2.5 degrees (fig. 2-34). The reflectors are mounted on three-axis pedestals and provide autotracking of a beacon or communication signal by conical scanning

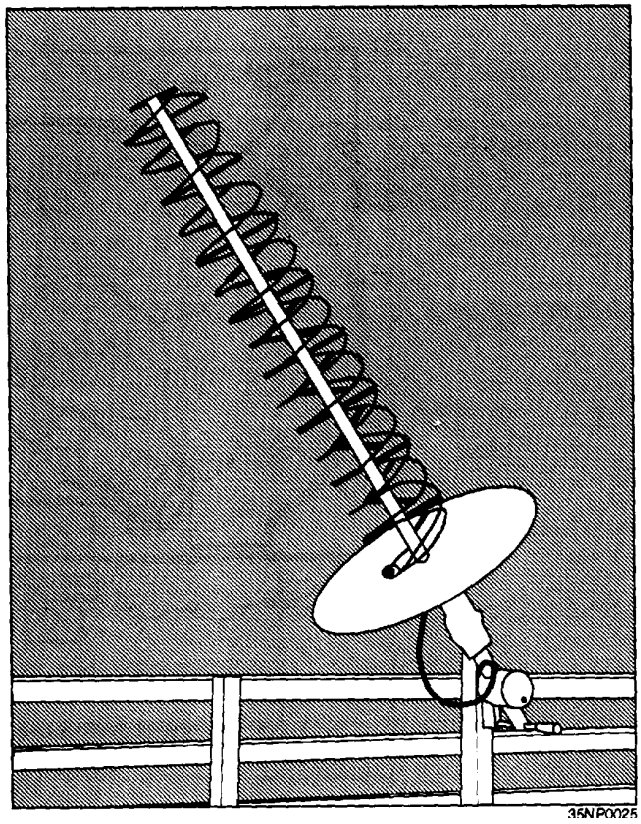


Figure 2-33.—Andrew 58622 shore antenna.

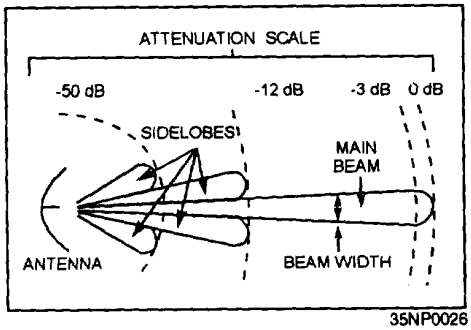


Figure 2-34.—AN/WSC-6(V) attenuation scale.

techniques. The antennas are radome enclosed and include various electronic components. Both a 7-foot model (fig. 2-35) and a 4-foot model (fig. 2-36) are operational in the fleet.

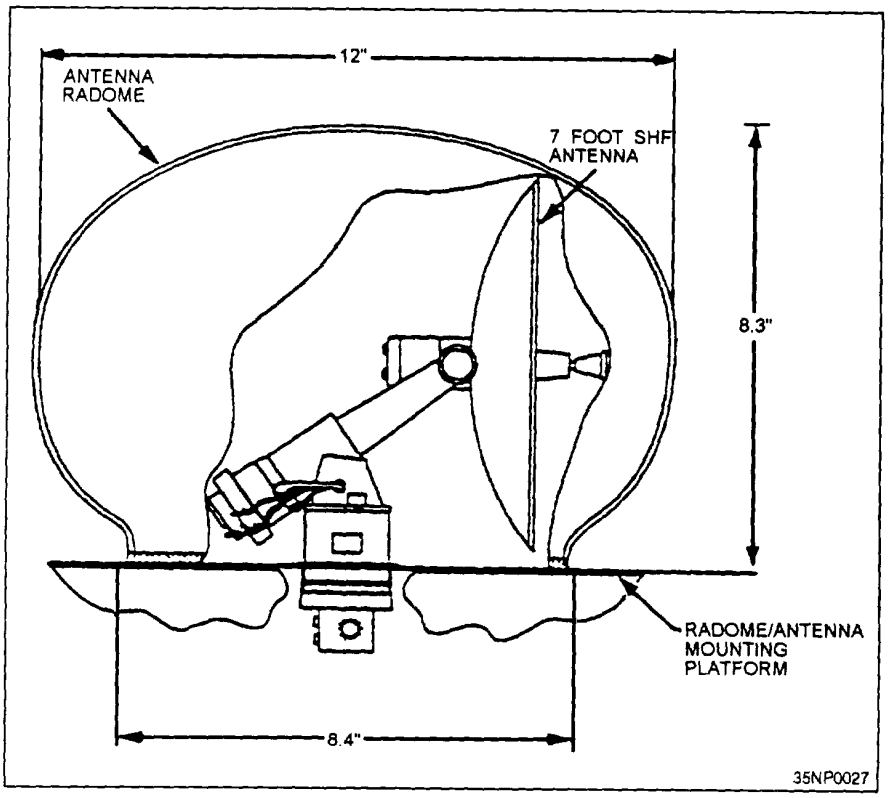


Figure 2-35.—Seven-foot shf SATCOM antenna.

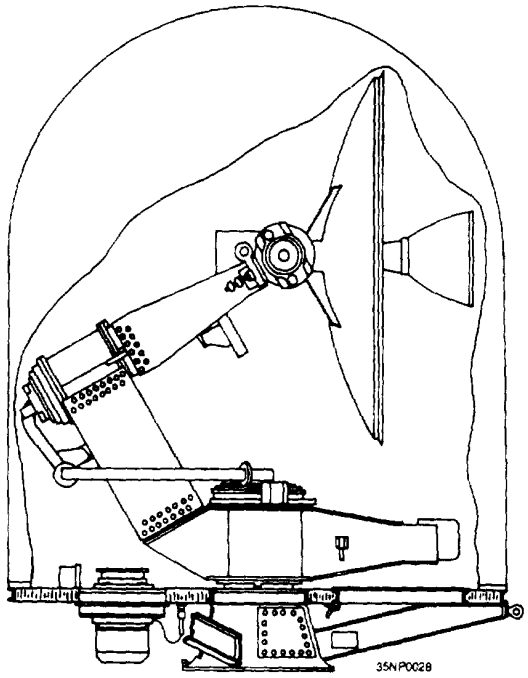


Figure 2-36.—Four-foot shf SATCOM antenna.

MATCHING NETWORKS

An antenna matching network consists of one or more parts (such as coils, capacitors, and lengths of transmission line) connected in series or parallel with the transmission line to reduce the standing wave ratio on the line. Matching networks are usually adjusted when they are installed and require no further adjustment for proper operation. Figure 2-37 shows a matching network outside of the antenna feedbox, with a sample matching network schematic.

Matching networks can also be built with variable components so they can be used for impedance matching over a range of frequencies. These networks are called antenna tuners.

Antenna tuners are usually adjusted automatically or manually each time the operating frequency is changed. Standard tuners are made with integral enclosures. Installation consists simply of mounting

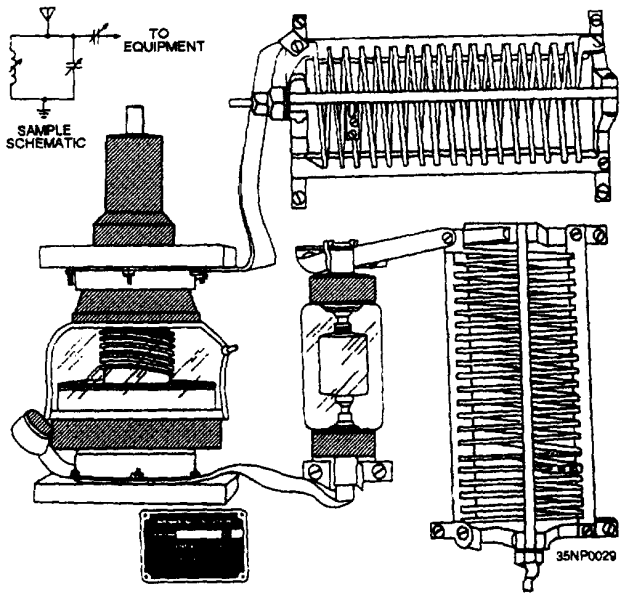


Figure 2-37.—Matching network.

the tuner, assembling the connections with the antenna and transmission line, and pressurizing the tuner, if necessary. Access must be provided to the pressure gauge and pressurizing and purging connections.

ANTENNA TUNING

For every frequency in the frequency spectrum, there is an antenna that is perfect for radiating at that frequency. By that we mean that all of the power being transmitted from the transmitter to the antenna will be radiated into space. Unfortunately, this is the ideal and not the rule. Normally, some power is lost between the transmitter and the antenna. This power loss is the result of the antenna not having the perfect dimensions and size to radiate perfectly all of the power delivered to it from the transmitter. Naturally, it would be unrealistic to carry a separate antenna for every frequency that a communications center is capable of radiating; a ship would have to have millions of antennas on board, and that would be impossible.

To overcome this problem, we use ANTENNA TUNING to lengthen and shorten antennas electrically to better match the frequency on which we want to transmit. The rf tuner is connected electrically to the antenna and is used to adjust the apparent physical length of the antenna by electrical means. This simply means that the antenna does not physically change length; instead, it is adapted electrically to the output frequency of the transmitter and "appears" to change its physical length. Antenna tuning is done by using antenna couplers, tuners, and multicouplers.

Antenna couplers and tuners are used to match a single transmitter or receiver to one antenna whereas antenna multicouplers are used to match more than one transmitter or receiver to one antenna for simultaneous operation. Some of the many antenna couplers that are in present use are addressed in the following paragraphs. For specific information on a particular coupler, refer to the appropriate equipment technical manual.

Antenna Coupler Group AN/URA-38

Antenna Coupler Group AN/URA-38 is an automatic antenna tuning system intended primarily for use with the AN/URT-23(V) operating in the high-frequency range. The equipment also includes provisions for manual and semiautomatic tuning, making the system readily adaptable for use with other radio transmitters. The manual tuning feature is useful when a failure occurs in the automatic tuning circuitry. Tuning can also be done without the use of rf power (silent tuning). This method is useful in installations where radio silence must be maintained except for brief transmission periods.

The antenna coupler matches the impedance of a 15-, 25-, 28-, or 35-foot whip antenna to a 50-ohm transmission line, at any frequency in the 2-to 30-MHz range. When the coupler is used with the AN/URT-23(V), control signals from the associated antenna coupler control unit automatically tune the coupler's matching network in less than 5 seconds. During manual and silent operation, the operator uses the controls mounted on the antenna coupler control unit to tune the antenna. A low-power (less than 250 watts) cw signal is required for tuning. Once tuned, the CU 938A/URA-38 is capable of handling 1000 watts PEP.

Antenna Coupler Groups AN/SRA-56, -57, and -58

Antenna coupler groups AN/SRA-56, -57, and -58 are designed primarily for shipboard use. Each

coupler group permits several transmitters to operate simultaneously into a single, associated, broadband antenna, thus reducing the total number of antennas required in the limited space aboard ship.

These antenna coupler groups provide a coupling path of prescribed efficiency between each transmitter and the associated antenna. They also provide isolation between transmitters, tunable bandpass filters to suppress harmonic and spurious transmitter outputs, and matching networks to reduce antenna impedances.

The three antenna coupler groups (AN/SRA-56, -57, -58) are similar in appearance and function, but they differ in frequency ranges. Antenna coupler group AN/SRA-56 operates in the 2- to 6-MHz frequency range. The AN/SRA-57 operates from 4 to 12 MHz, and the AN/SRA-58 operates from 10 to 30 MHz. When more than one coupler is used in the same frequency range, a 15 percent frequency separation must be maintained to avoid any interference.

Antenna Coupler Group AN/SRA-33

Antenna coupler group AN/SRA-33 operates in the uhf (225-400 Mhz) frequency range. It provides isolation between as many as four transmitter and receiver combinations operating simultaneously into a common uhf antenna without degrading operation. The AN/SRA-33 is designed for operation with shipboard radio set AN/WSC-3. The AN/SRA-33 consists of four antenna couplers (CU-1131/SRA-33 through CU-1134/SRA-33), a control power supply C-4586/SRA-33, an electronic equipment cabinet CY-3852/SRA-33, and a set of special-purpose cables.

OA-9123/SRC

The OA-9123/SRC multicoupler enables up to four uhf transceivers, transmitters, or receivers to operate on a common antenna. The multicoupler provides low insertion loss and highly selective filtering in each of the four ports. The unit is interface compatible with the channel select control signals from radio sets AN/WSC-3(V) (except (V)1). The unit is self-contained and is configured to fit into a standard 19-inch open equipment rack.

The OA-9123/SRC consists of a cabinet assembly, control power supply assembly, and four identical filter assemblies. This multicoupler is a state-of-the-art replacement for the AN/SRA-33 and only requires about half of the space.

RECEIVING ANTENNA DISTRIBUTION SYSTEMS

Receiving antenna distribution systems operate at low power levels and are designed to prevent multiple signals from being received. The basic distribution system has several antenna transmission lines and several receivers, as shown in figure 2-38. The system includes two basic patch panels, one that terminates the antenna transmission lines, and the other that terminates the lines leading to the receivers. Thus, any antenna can be patched to any receiver via patch cords.

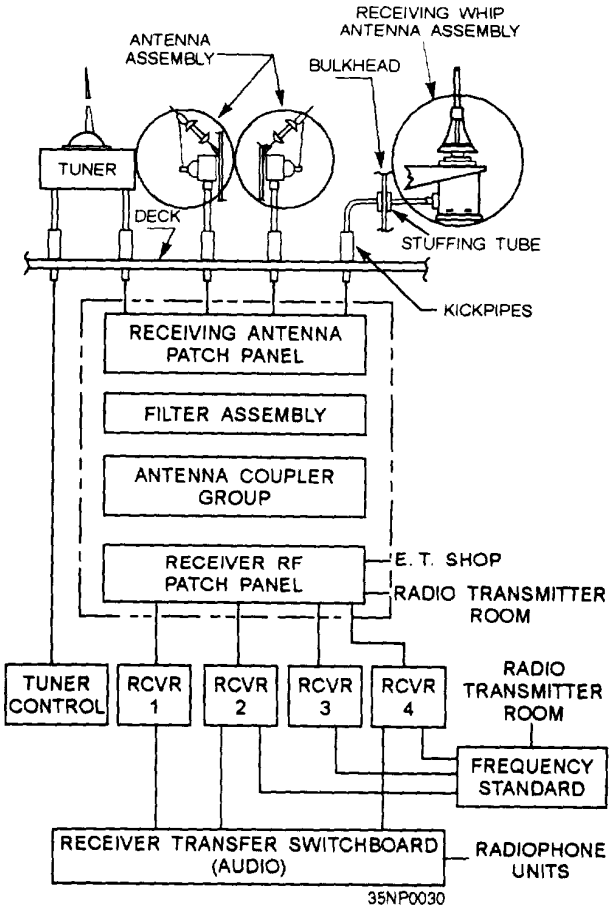


Figure 2-38.—Receive signal distribution system.

Some distribution systems will be more complex. That is, four antennas can be patched to four receivers, or one antenna can be patched to more than one receiver via the multicouplers.

RECEIVING MULTICOUPLER AN/SRA-12

The AN/SRA-12 filter assembly multicoupler provides seven radio frequency channels in the 14-kHz to 32-MHz frequency range. Any of these channels may be used independently of the other channels, or they may operate simultaneously. Connections to the receiver are made by coaxial patch cords, which are short lengths of cable with a plug attached to each end.

ANTENNA COUPLER GROUPS AN/SRA-38, AN/SRA-39, AN/SRA-40, AN/SRA-49, AN/SRA-49A, and AN/SRA-50

These groups are designed to connect up to 20 mf and hf receivers to a single antenna, with a highly selective degree of frequency isolation. Each of the six coupler groups consists of 14 to 20 individual antenna couplers and a single-power supply module, all are slide-mounted in a special electronic equipment rack. An antenna input distribution line termination (dummy load) is also supplied. In addition, there are provisions for patching the outputs from the various antenna couplers to external receivers.

RADAR ANTENNAS

Radar antennas are usually directional antennas that radiate energy in one lobe or beam. The two most important characteristics of directional antennas are directivity and power gain. Most radar systems use parabolic antennas. These antennas use parabolic reflectors in different variations to focus the radiated energy into a desired beam pattern.

While most radar antennas are parabolic, other types such as the corner reflector, the broadside array, and horn radiators may also be used.

PARABOLIC REFLECTORS

To understand why parabolic reflectors are used for most radar antennas, you need to understand how radio waves behave. A point source, such as an omnidirectional antenna produces a spherical radiation pattern, or spherical wavefront. As the sphere expands, the energy contained in a given surface area decreases rapidly. At a relatively short distance from the antenna, the energy level is so small that its reflection from a target would be useless in a radar system.

A solution to this problem is to form the energy into a PLANE wavefront, In a plane wavefront, all of the energy travels in the same direction, thus providing more energy to reflect off of a target. To concentrate the energy even further, a parabolic reflector is used to shape the plane wavefront's energy into a beam of energy. This concentration of energy provides a maximum amount of energy to be reflected off of a target, making detection of the target much more probable.

How does the parabolic reflector focus the radio waves? Radio waves behave much as light waves do. Microwaves travel in straight lines as do light rays. They may be focused or reflected, just as light rays may be. In figure 2-39, a point-radiation source is placed at the focal point F. The field leaves this antema with a spherical wavefront. As each part of the wavefront moving toward the reflector reaches the reflecting surface, it is shifted 180 degrees in phase and sent outward at angles that cause all parts of the field to travel in parallel paths. Because of the shape of a parabolic surface, all paths from F to the reflector and back to line XY are the same length. Therefore, all parts of the field arrive at line XY at the same time after reflection.

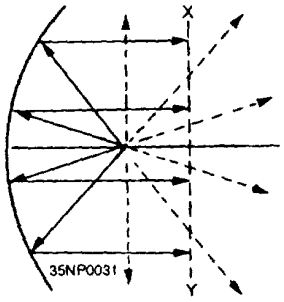


Figure 2-39.—Parabolic reflector radiation.

Energy that is not directed toward the paraboloid (dotted lines in fig. 2-39) has a wide-beam characteris-

tic that would destroy the narrow pattern from the parabolic reflector. This destruction is prevented by the use of a hemispherical shield (not shown) that directs most of what would otherwise be spherical radiation toward the parabolic surface. Without the shield, some of the radiated field would leave the radiator directly, would not be reflected, and would serve no useful purpose. The shield makes the beamsharper, and concentrates the majority of the power in the beam. The same results can be obtained by using either a parasitic array to direct the radiated field back to the reflector, or a feed horn pointed at the paraboloid.

The radiation pattern of the paraboloid contains a major lobe, which is directed along the axis of the paraboloid, and several minor lobes, as shown in figure 2-40. Very narrow beams are possible with this type of reflector. View A of figure 2-41 illustrates the parabolic reflector.

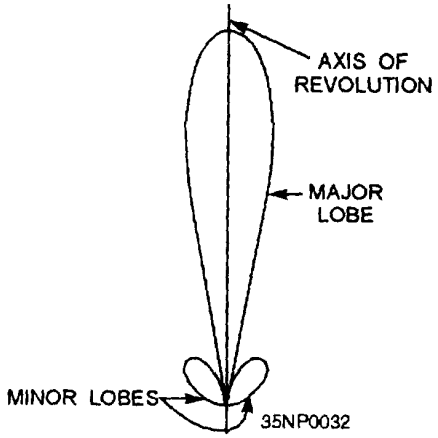


Figure 2-40.—Parabolic radiation pattern.

Truncated Paraboloid

While the complete parabolic reflector produces a pencil-shaped beam, partial parabolic reflectors produce differently shaped beams. View B of figure 2-41 shows a horizontally truncated, or vertically shortened, paraboloid. This type of reflector is designed to produce a beam that is narrow horizontally but wide vertically. Since the beam is wide vertically, it will detect aircraft at different altitudes without changing the tilt of the antenna. It also works well for surface search radars to overcome the pitch and roll of the ship.

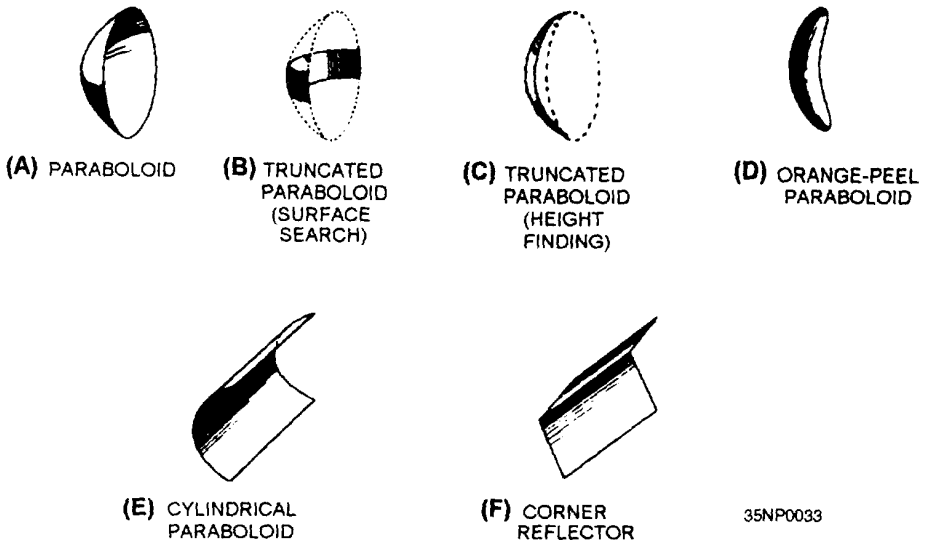


Figure 2-41.—Reflector shapes.

The truncated paraboloid reflector may be used in height-finding systems if the reflector is rotated 90 degrees, as shown in view C of figure 2-41. This type of reflector produces a beam that is wide horizontally but narrow vertically. The beam pattern is spread like a horizontal fan. Such a fan-shaped beam can be used to determine elevation very accurately.

Orange-Peel Paraboloid

A section of a complete circular paraboloid, often called an ORANGE-PEEL REFLECTOR because of its shape, is shown in view D of figure 2-41. Since the reflector is narrow in the horizontal plane and wide in the vertical, it produces a beam that is wide in the horizontal plane and narrow in the vertical. In shape, the beam resembles a huge beaver tail. This type of antenna system is generally used in height-finding equipment.

Cylindrical Paraboloid

When a beam of radiated energy noticeably wider in one cross-sectional dimension than in the other is desired, a cylindrical paraboloid section approximating a rectangle can be used. View E of figure 2-41 illustrates this antenna. A parabolic cross section is in one dimension only; therefore, the reflector is directive in one plane only. The cylindrical paraboloid reflector is either fed by a linear array of dipoles, a slit in the side of a waveguide, or by a thin waveguide radiator. Rather than a single focal point, this type of reflector has a series of focal points forming a straight line. Placing the radiator, or radiators, along this focal line produces a directed beam of energy. As the width of the parabolic section is changed, different beam shapes are produced. This type of antenna system is used in search systems and in ground control approach (gca) systems.

CORNER REFLECTOR

The corner-reflector antenna consists of two flat conducting sheets that meet at an angle to form a corner, as shown in view F of figure 2-41. This reflector is normally driven by a half-wave radiator located on a line that bisects the angle formed by the sheet reflectors.

BROADSIDE ARRAY

Desired beam widths are provided for some vhf radars by a broadside array, such as the one shown in figure 2-42. The broadside array consists of two or more half-wave dipole elements and a flat reflector. The elements are placed one-half wavelength apart and parallel to each other. Because they are excited in phase, most of the radiation is perpendicular or broadside to the plane of the elements. The flat reflector is located approximately one-eighth wavelength behind the dipole elements and makes possible the unidirectional characteristics of the antenna system.

HORN RADIATORS

Horn radiators, like parabolic reflectors, may be used to obtain directive radiation at microwave frequencies. Because they do not involve resonant elements, horns have the advantage of being usable over a wide frequency band.

The operation of a horn as an electromagnetic directing device is analogous to that of acoustic horns. However, the throat of an acoustic horn usually has dimensions much smaller than the sound wavelengths for which it is used, while the throat of the electromagnetic horn has dimensions that are comparable to the wavelength being used.

Horn radiators are readily adaptable for use with waveguides because they serve both as an impedance-

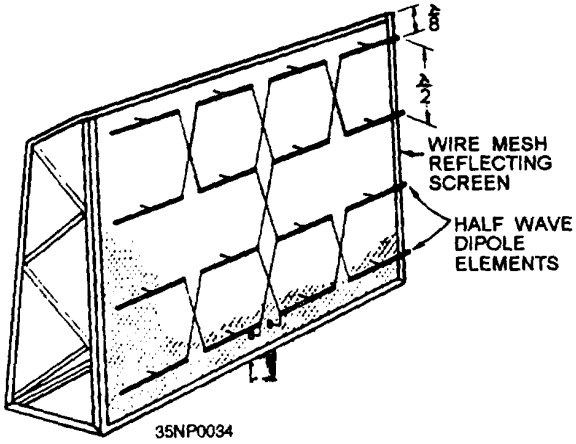


Figure 2-42.—Broadside array.

matching device and as a directional radiator. Horn radiators may be fed by coaxial or other types of lines.

Horns are constructed in a variety of shapes as illustrated in figure 2-43. The shape of the horn and the dimensions of the length and mouth largely determine the field-pattern shape. The ratio of the horn length to mouth opening size determines the beam angle and, thus, the directivity. In general, the larger the opening of the horn, the more directive is the resulting field pattern.

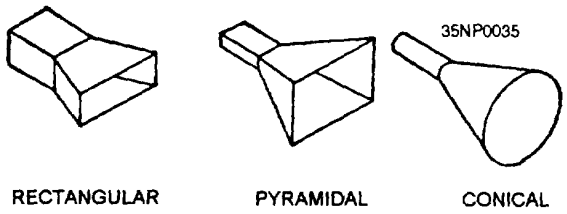


Figure 2-43.—Horn radiators.

FEEDHORNS

A waveguide horn, called a FEEDHORN, may be used to feed energy into a parabolic dish. The directivity of this feedhorn is added to that of the parabolic dish. The resulting pattern is a very narrow and concentrated beam. In most radars, the feedhorn is covered with a window of polystyrene fiberglass to prevent moisture and dirt from entering the open end of the waveguide.

One problem associated with feedhorns is the SHADOW introduced by the feedhorn if it is in the path of the beam. (The shadow is a dead spot directly in front of the feedhorn.) To solve this problem the feedhorn can be offset from center. This location change takes the feedhorn out of the path of the rf beam and eliminates the shadow. An offset feedhorn is shown in figure 2-44.

RADAR SYSTEMS

Now that you have a basic understanding of how radar antennas operate, we will introduce you to a few of the radar systems currently in use.

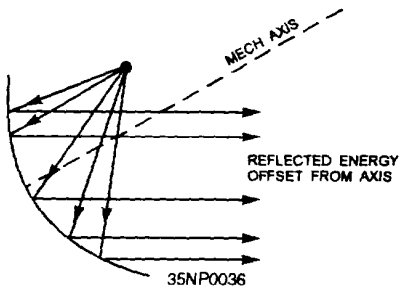


Figure 2-44.—Offset feedhorn.

AN/GPN-27(ASR-8) AIR SURVEILLANCE RADAR

The AN/GPN-27(ASR-8) (fig. 2-45) antenna radiates a beam 1.5 degrees in azimuth and shaped in elevation to produce coverage of up to approximately 32 degrees above the horizon. This provides a maplike presentation of aircraft within 55 nautical miles of an airport terminal. The antenna azimuth

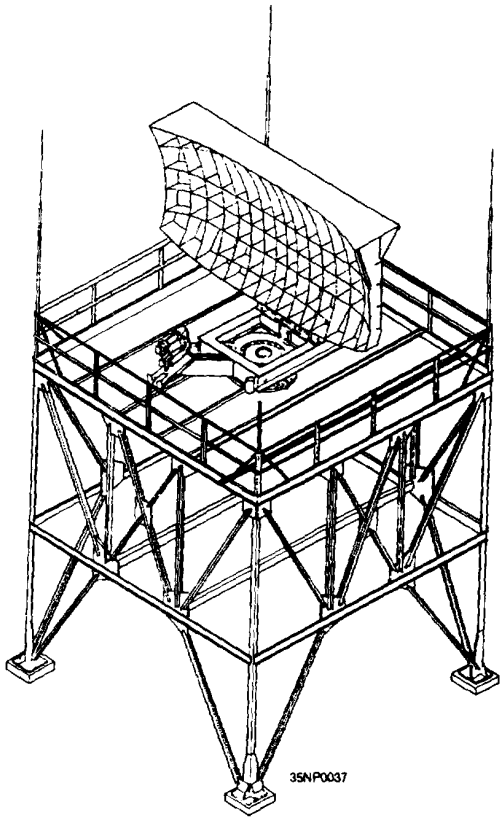


Figure 2-45.—AN/GPN-27(ASR-8) air surveillance radar.

pulse generator (APG), located in the rotary joint, transmits to the radar indicator azimuth information corresponding to beam direction. Polarization of the radiated energy can be remotely switched to either linear or circular polarization. The reflector has a modified parabolic shape designed to produce an approximately cosecant squared beam in the elevation plane. The reflector surface, covered with expanded aluminum screen, is 16.1 feet wide and 9 feet high. The antenna feedhorn, which mounts on the polarizer, provides impedance matching between the waveguide system and free space, and produces the desired feed pattern to illuminate the reflector. A radome over the horn aperture excludes moisture and foreign matter, and provides a pressure seal.

AS-3263/SPS-49(V)

The AS-3263/SPS-49(V) antenna (fig. 2-46) consists of three major sections: the antenna base and pedestal assembly, the feedhorn and feedhorn support boom, and the reflector assembly.

The base assembly provides a surface for mounting the antenna to the ship. It also contains the azimuth drive gearbox. The gearbox is driven by the azimuth drive motor, which drives the pedestal in azimuth through a pinion gear mated to a ring gear located at the bottom of the cone-shaped pedestal assembly, The azimuth drive circuits rotate the antenna through 360 degrees at speeds of 6 rpm and 12 rpm.

The reflector and the feedhorn support boom are mounted on a trunnion, allowing the elevation angle of the rf beam to be controlled by a jackscrew located behind the reflector. The jackscrew is rotated by the elevation drive gearbox, which is connected to two dc motors. The rf energy to the feedhorn is routed through elevation and azimuth rotary joints located within the pedestal.

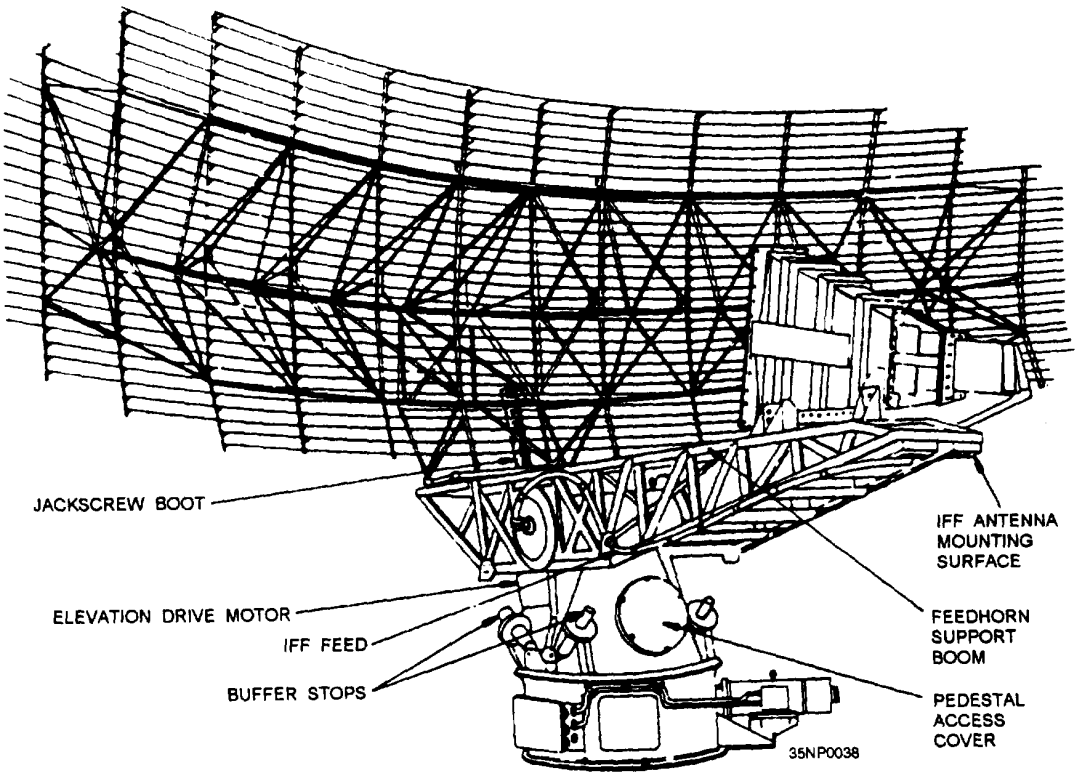


Figure 2-46.—AS-3263/SPS-49(V) antenna.

The reflector is 24 feet wide and has a double-curved surface composed of a series of horizontal members that form a reflecting surface for the horizontally polarized C-band energy. The antenna has a 28-dB gain, with a beamwidth of 9 degrees minimum vertically and approximately 3.3 degrees horizontally. Antenna roll and pitch stabilization limits are plus or minus 25 degrees, Stabilization accuracy is plus or minus 1 degree with the horizontal plane.

The antenna is equipped with a safety switch located near the antenna pedestal area. The safety switch disables the azimuth and elevation functions in the antenna and the radiate function in the transmitter to provide protection for personnel working on the antenna.

OE-172/SPS-55

The OE-172/SPS-55 antenna group consists of the antenna and the antenna pedestal. The antenna group is mast-mounted by means of four bolt holes on the base of the pedestal.

The antenna consists of two waveguide slotted arrays mounted back-to-back. One array provides linear polarization, while the other provides circular polarization. The array used is selected by means of a remotely controlled waveguide switch located on the pedestal. Linear polarization is used for most conditions. Circular polarization is used to reduce return echoes from precipitation. Each antenna forms a fan beam that is narrow in the azimuth plane and moderately broad in the elevation plane.

Figure 2-47 shows a cross-section of the SPS-55 antenna. During transmission, the rf signal enters the antenna through a feed waveguide and then enters a feed manifold region of 80 periodic narrow-wall slots. The slots are skewed in angle and alternated in direction of skew. They are separated by approximately one-half wavelength, resulting in broadside radiation into the sectoral horn region of the antenna. The horizontally polarized radiation from the manifold travels in the horn region toward the aperture, where it encounters an array of vertical sheet metal slats.

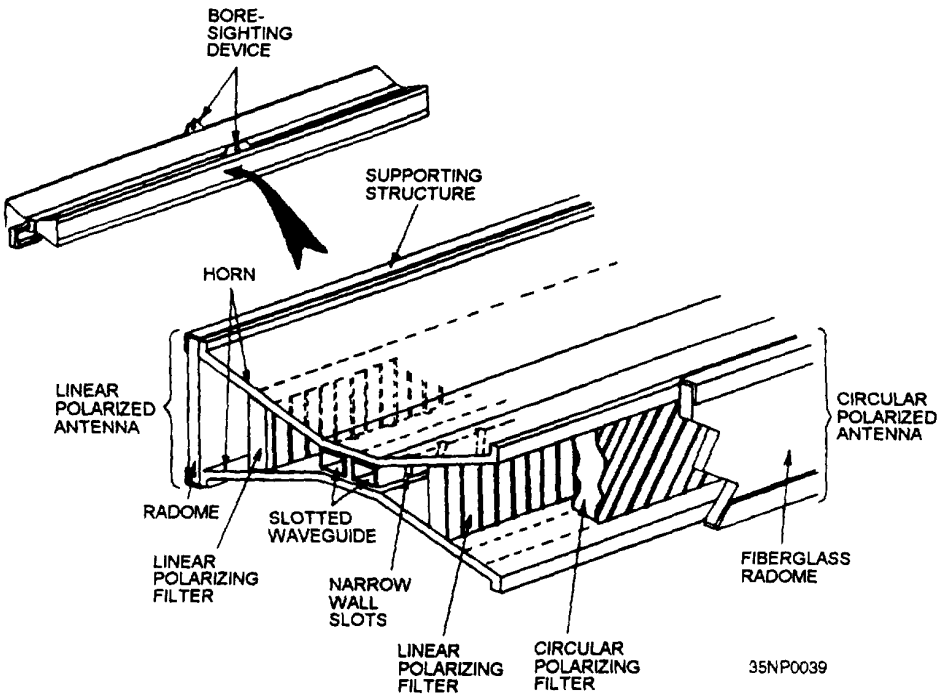


Figure 2-47.—SPS-55 antenna cross section.

This array is a polarizing filter, which ensures that only horizontally polarized energy travels from the horn region. The antenna scans at a rate of 16 rpm and produces an absolute gain of 31 dB at midband.

AN/SPN-35A AIRCRAFT CONTROL APPROACH RADAR

The AN/SPN-35A (fig 2-48) is a carrier-controlled-approach (CCA) radar set used for precision landing approaches during adverse weather conditions. The radar displays both azimuth and elevation data, which enables the radar operator to direct aircraft along a predetermined glide path and azimuth course line to a transition point approximately 2 miles from the ramp of the flight deck.

The azimuth antenna, AS-1292/TPN-8, functions in the azimuth rf line for radiation and reception of X-band rf pulses. The azimuth antenna comprises a truncated paraboloid-type reflector with an offset feedhorn and a polarizer assembly that provides remote-controlled selection of either horizontal or circular polarization. The antenna is located above the azimuth drive assembly on the stabilized yoke. The azimuth drive can rotate the antenna in either 360 degrees or in limited-sector modes of operation in the horizontal plane.

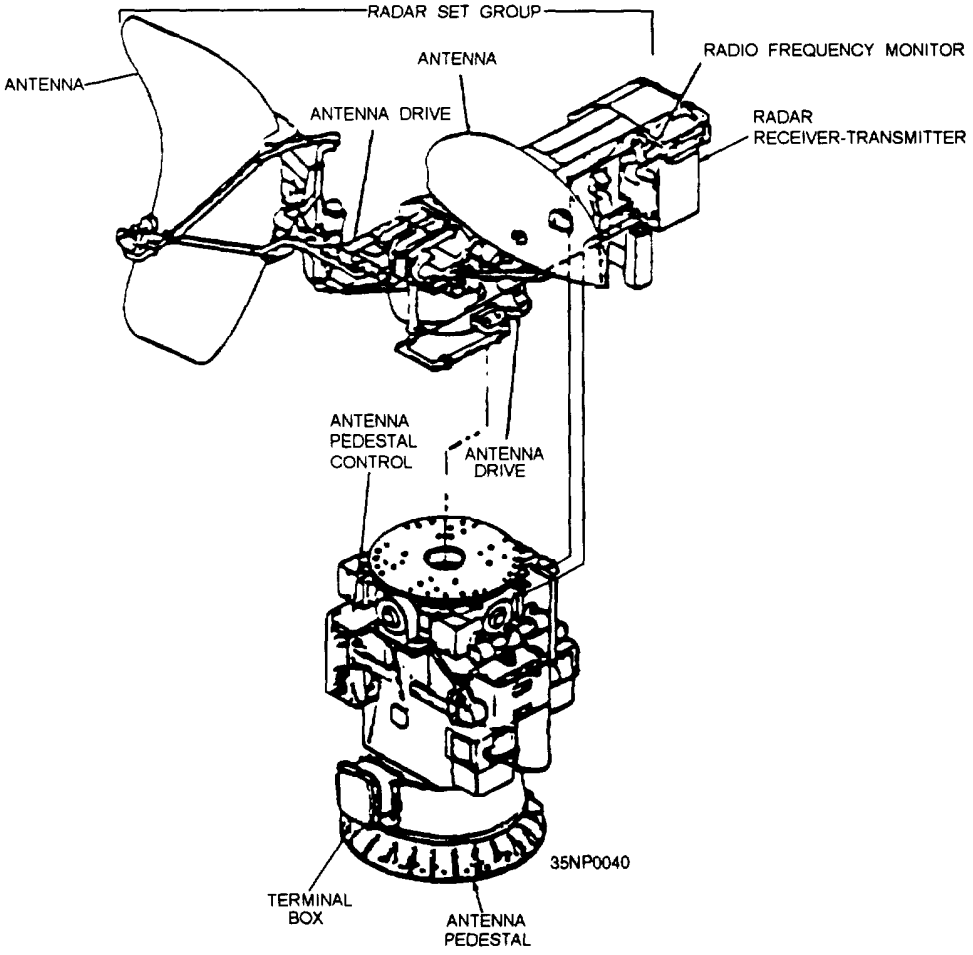


Figure 2-48.—AN/SPN-3SA aircraft control approach radar.

The elevation antenna, AS-1669/SPN-35, is a truncated paraboloid-type reflector with a dual-channel feedhorn and a polarizer assembly providing monopulse-type radiation and reception of X-band rf pulses. The horizontal shape of the laminated fiberglass reflector is cosecanted. The dual-channel feedhorn and polarizer are fixed in circular polarization by an external grid device. The elevation antenna is stabilized-yoke mounted on the elevation drive assembly adjacent to the azimuth antenna. The elevation drive provides the required motion for the elevation antenna and locks electrically with the search drive when the radar set operates in the precision mode.

The radar operates in three modes, final, surveillance, and simultaneous, with each antenna acting independently. In the final (precision) mode, the azimuth antenna scans a 30-degree sector (60-degree sector optional) while the elevation antenna scans a 10-degree sector (35-degree sector optional). In the surveillance mode the azimuth antenna rotates through the full 360-degree search pattern at 16 rpm while the elevation antenna scans a 10-degree sector. In the simultaneous mode, the azimuth antenna rotates through the full 360-degrees search pattern in 60-degree increments while the elevation antenna scans a 10-degree sector. The data rate in this mode is approximately 16 azimuth sweeps and 24 elevation sweeps every 60 seconds.

The antenna pedestal control stabilizes the azimuth and elevation antennas for plus or minus 3 degrees of pitch and plus or minus 10 degrees of roll.

RF SAFETY PRECAUTIONS

Although radio frequency emissions are usually harmless, there are still certain safety precautions you should follow whenever you are near high-power rf sources. Normally, electromagnetic radiation from transmission lines and antennas isn't strong enough to electrocute personnel. However, it may lead to other accidents and can compound injuries. Voltages may be induced into metal objects both above and below ground, such as wire guys, wire cable, hand rails, and ladders. If you come into contact with these objects, you may receive a shock or an rf burn. The shock can cause you to jump involuntarily, to fall into nearby equipment, or, when working aloft, to fall from the elevated work area. Take care to ensure that all transmission lines or antennas are de-energized before working on or near them.

When working aloft aboard ship, be sure to use a working aloft chit. This will ensure that all radiators, not only those on your own ship but also those nearby are secured while you are aloft.

ALWAYS obey rf radiation warning signs and keep a safe distance from radiating antennas. The six types of warning signs for rf radiation hazards are shown in figure 2-49.

The two primary safety concerns associated with rf fields are rf burns and injuries caused by dielectric heating.

RF BURNS

Close or direct contact with rf transmission lines or antennas may result in rf burns caused by induced voltages. These burns are usually deep, penetrating, third-degree burns. To heal properly, rf burns must heal from the inside toward the skin's surface. Do **NOT** take rf burns lightly. To prevent infection, you must give proper attention to **ALL** rf burns, including the small pinhole burns. **ALWAYS** seek treatment for any rf burn or shock and report the incident to your supervisor so appropriate action can be taken to correct the hazard.

DIELECTRIC HEATING

While the severity of rf burns may vary from minor to major, burns or other damage done by DIELECTRIC HEATING may result in long-term injury, or even death. Dielectric heating is the heating of an insulating material caused by placing it in a high-frequency electric field. The heat results from the rapid reversal of molecular polarization dielectric material.

When a human is in an rf field, the body acts as the dielectric. If the power in the rf field exceeds 10 milliwatts per centimeter, the individual will have a noticeable rise in body temperature. Basically, the body is “cooking” in the rf field. The vital organs

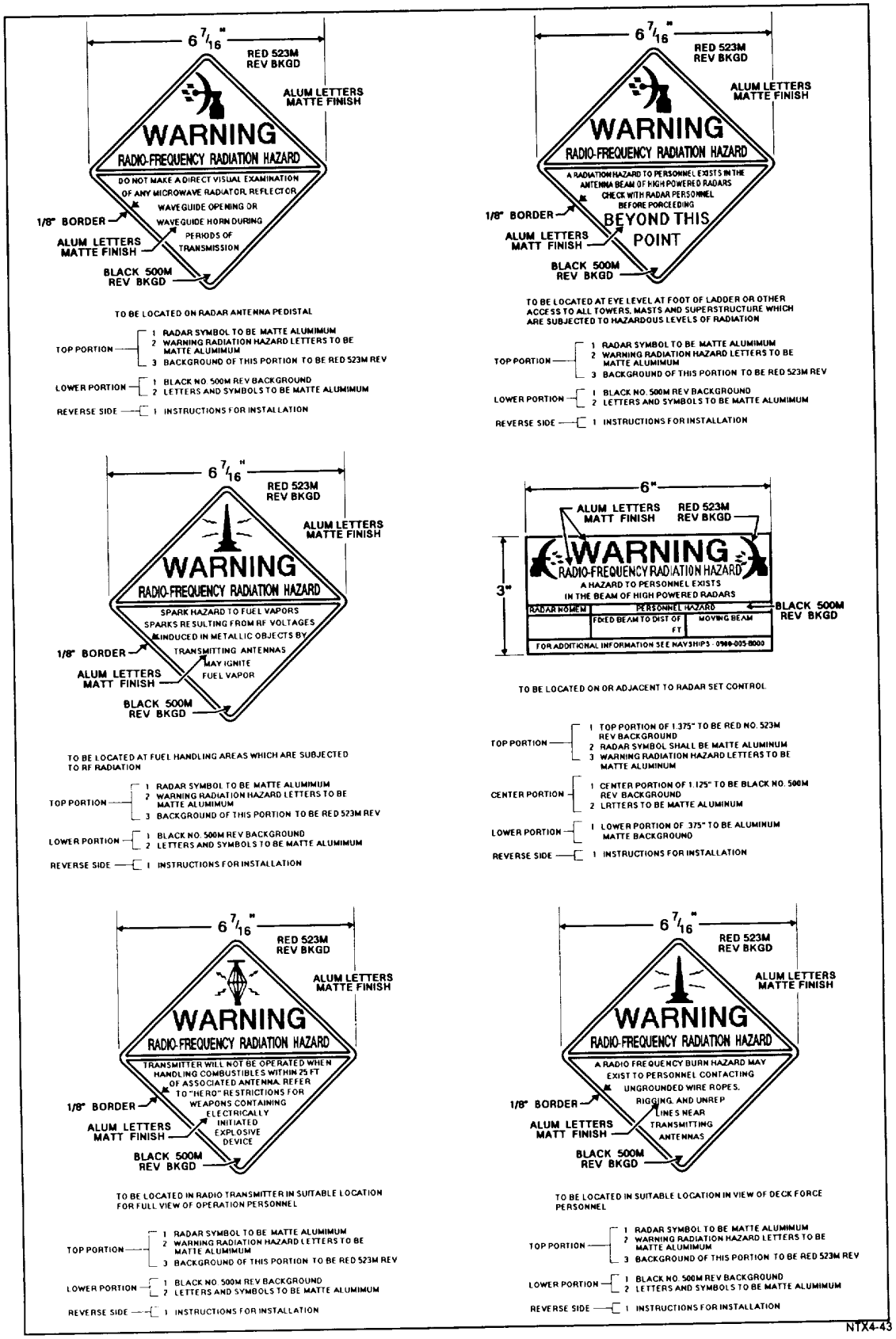


Figure 2-49.—Rf radiation warning signs

of the body are highly susceptible to dielectric heating. The eyes are also highly susceptible to dielectric heating. Do NOT look directly into devices radiating rf energy. Remember, rf radiation can be dangerous. For your own safety, you must NOT stand directly in the path of rf radiating devices.

PRECAUTIONS WHEN WORKING ALOFT

As we mentioned earlier, it is extremely important to follow all safety precautions when working aloft. Before you work on an antenna, ensure that it is tagged out properly at the switchboard to prevent it from becoming operational. Always be sure to secure the motor safety switches for rotating antennas. Have the switches tagged and locked open before you begin working on or near the antenna.

When working near a stack, draw and wear the recommended oxygen breathing apparatus. Among other toxic substances, stack gas contains carbon monoxide. Carbon monoxide is too unstable to build up to a high concentration in the open, but prolonged exposure to even small quantities is dangerous.

For more detailed information concerning the dangers and hazards of rf radiation, refer to the NAVELEX technical manual, *Electromagnetic Radiation Hazards.* NAVELEX 0967-LP-624-6010.

This completes chapter 2. In chapter 3, we will discuss transmission lines and waveguides.

CHAPTER 3

INTRODUCTION TO TRANSMISSION LINES AND WAVEGUIDES

A TRANSMISSION LINE is a device designed to guide electrical energy from one point to another. It is used, for example, to transfer the output rf energy of a transmitter to an antenna. This energy will not travel through normal electrical wire without great losses. Although the antenna can be connected directly to the transmitter, the antenna is usually located some distance away from the transmitter. On board ship, the transmitter is located inside a radio room, and its associated antenna is mounted on a mast. A transmission line is used to connect the transmitter and the antenna.

The transmission line has a single purpose for both the transmitter and the antenna. This purpose is to transfer the energy output of the transmitter to the antenna with the least possible power loss. How well this is done depends on the special physical and electrical characteristics (impedance and resistance) of the transmission line.

TRANSMISSION LINE THEORY

The electrical characteristics of a two-wire transmission line depend primarily on the construction of the line. The two-wire line acts like a long capacitor. The change of its capacitive reactance is noticeable as the frequency applied to it is changed. Since the long conductors have a magnetic field about them when electrical energy is being passed through them, they also exhibit the properties of inductance. The values of inductance and capacitance presented depend on the various physical factors that we discussed earlier. For example, the type of line used, the dielectric in the line, and the length of the line must be considered. The effects of the inductive and capacitive reactance of the line depend on the frequency applied. Since no dielectric is perfect, electrons manage to move from one conductor to the other through the dielectric. Each type of two-wire transmission line also has a conductance value. This conductance value represents the value of the current flow that may be expected through the insulation, If the line is uniform (all values equal at each unit length), then one small section of the line may represent several feet. This illustration of a two-wire transmission line will be used throughout the discussion of transmission lines; but, keep in mind that the principles presented apply to all transmission lines. We will explain the theories using LUMPED CONSTANTS and DISTRIBUTED CONSTANTS to further simplify these principles.

LUMPED CONSTANTS

A transmission line has the properties of inductance, capacitance, and resistance just as the more conventional circuits have. Usually, however, the constants in conventional circuits are lumped into a single device or component. For example, a coil of wire has the property of inductance. When a certain amount of inductance is needed in a circuit, a coil of the proper dimensions is inserted. The inductance of the circuit is lumped into the one component. Two metal plates separated by a small space, can be used to supply the required capacitance for a circuit. In such a case, most of the capacitance of the circuit is lumped into this one component. Similarly, a fixed resistor can be used to supply a certain value of circuit resistance as a lumped sum. Ideally, a transmission line would also have its constants of inductance, capacitance, and resistance lumped together, as shown in figure 3-1. Unfortunately, this is not the case. Transmission line constants are as described in the following paragraphs.

DISTRIBUTED CONSTANTS

Transmission line constants, called distributed constants, are spread along the entire length of the transmission line and cannot be distinguished separately. The amount of inductance, capacitance, and resistance depends on the length of the line, the size of the conducting wires, the spacing between the

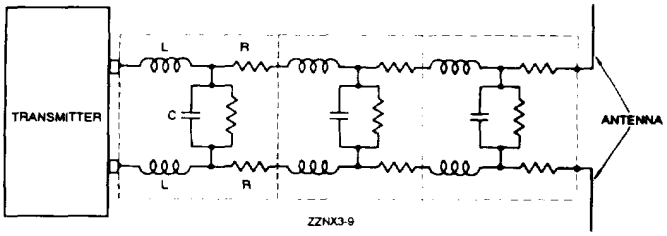


Figure 3-1.—Two-wire transmission line.

wires, and the dielectric (air or insulating medium) between the wires. The following paragraphs will be useful to you as you study distributed constants on a transmission line.

Inductance of a Transmission Line

When current flows through a wire, magnetic lines of force are set up around the wire. As the current increases and decreases in amplitude, the field around the wire expands and collapses accordingly. The energy produced by the magnetic lines of force collapsing back into the wire tends to keep the current flowing in the same direction. This represents a certain amount of inductance, which is expressed in microhenrys per unit length. Figure 3-2 illustrates the inductance and magnetic fields of a transmission line.

Capacitance of a Transmission Line

Capacitance also exists between the transmission line wires, as illustrated in figure 3-3. Notice that the two parallel wires act as plates of a capacitor and that the air between them acts as a dielectric. The capacitance between the wires is usually expressed in picofarads per unit length. This electric field between the wires is similar to the field that exists between the two plates of a capacitor.

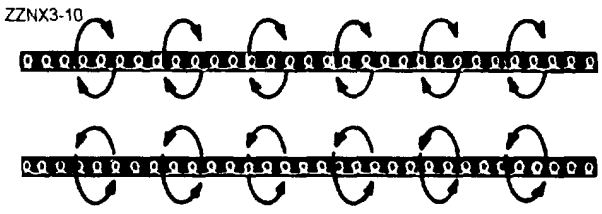


Figure 3-2.—Distributed inductance.

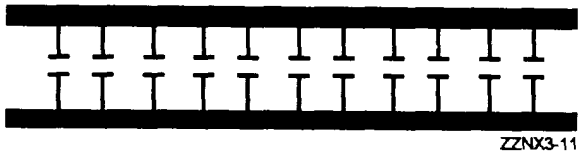


Figure 3-3.—Distributed capacitance.

Resistance of a Transmission Line

The transmission line shown in figure 3-4 has electrical resistance along its length. This resistance is usually expressed in ohms per unit length and is shown as existing continuously from one end of the line to the other.

Figure 3-4.—Distributed resistance.

Leakage Current

Since any dielectric, even air, is not a perfect insulator, a small current known as LEAKAGE CURRENT flows between the two wires. In effect, the insulator acts as a resistor, permitting current to pass between the two wires. Figure 3-5 shows this leakage path as resistors in parallel connected between the two lines. This property is called CONDUCTANCE (G) and is the opposite of resistance. Conductance in transmission lines is expressed as the reciprocal of resistance and is usually given in micromhos per unit length.

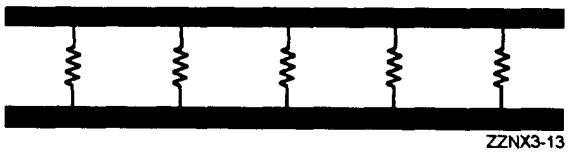


Figure 3-5.—Leakage in a transmission line.

ELECTROMAGNETIC FIELDS

The distributed constants of resistance, inductance, and capacitance are basic properties common to all transmission lines and exist whether or not any current flow exists. As soon as current flow and voltage exist in a transmission line, another property becomes quite evident. This is the presence of an electromagnetic field, or lines of force, about the wires of the transmission line. The lines of force themselves are not visible; however, understanding the force that an electron experiences while in the field of these lines is very important to your understanding of energy transmission.

There are two kinds of fields; one is associated with voltage and the other with current. The field associated with voltage is called the ELECTRIC (E) FIELD. It exerts a force on any electric charge placed in it. The field associated with current is called a MAGNETIC (H) FIELD, because it tends to exert a force on any magnetic pole placed in it. Figure 3-6 illustrates the way in which the E fields and H fields tend to orient themselves between conductors of a typical two-wire transmission line. The illustration shows a cross section of the transmission lines. The E field is represented by solid lines and the H field by dotted lines. The arrows indicate the direction of the lines of force. Both fields normally exist together and are spoken of collectively as the electromagnetic field.

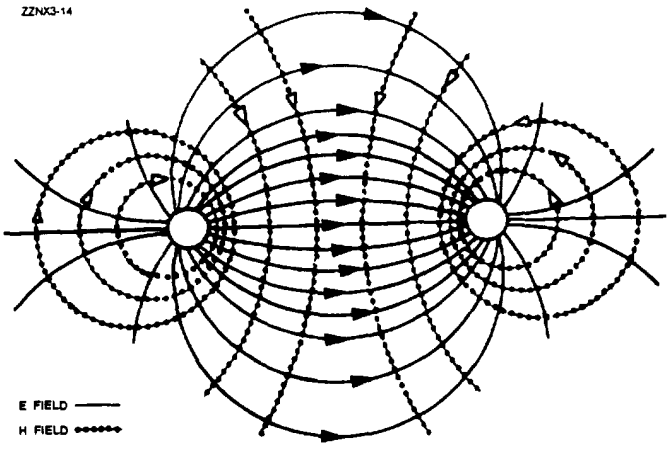


Figure 3-6.—Fields between conductors.

CHARACTERISTIC IMPEDANCE

You can describe a transmission line in terms of its impedance. The ratio of voltage to current (E_{in}/I_{in}) at the input end is known as the INPUT IMPEDANCE (Z_{in}). This is the impedance presented to the transmitter by the transmission line and its load, the antenna. The ratio of voltage to current at the output (E_{OUT}/I_{OUT}) end is known as the OUTPUT IMPEDANCE (Z_{OUT}). This is the impedance presented to the load by the transmission line and its source. If an infinitely long transmission line could be used, the ratio of voltage to current at any point on that transmission line would be some particular value of impedance. This impedance is known as the CHARACTERISTIC IMPEDANCE.

The maximum (and most efficient) transfer of electrical energy takes place when the source impedance is matched to the load impedance. This fact is very important in the study of transmission lines and antennas. If the characteristic impedance of the transmission line and the load impedance are equal, energy from the transmitter will travel down the transmission line to the antenna with no power loss caused by reflection.

LINE LOSSES

The discussion of transmission lines so far has not directly addressed LINE LOSSES; actually some losses occur in all lines. Line losses may be any of three types—COPPER, DIELECTRIC, and RADIATION or INDUCTION LOSSES.

NOTE: Transmission lines are sometimes referred to as rf lines. In this text the terms are used interchangeably.

Copper Losses

One type of copper loss is I^2R LOSS. In rf lines the resistance of the conductors is never equal to zero. Whenever current flows through one of these conductors, some energy is dissipated in the form of heat. This heat loss is a POWER LOSS. With copper braid, which has a resistance higher than solid tubing, this power loss is higher.

Another type of copper loss is due to SKIN EFFECT. When dc flows through a conductor, the movement of electrons through the conductor's cross section is uniform, The situation is somewhat different when ac is applied. The expanding and collapsing fields about each electron encircle other electrons. This phenomenon, called SELF INDUCTION, retards the movement of the encircled electrons. The flux density at the center is so great that electron movement at this point is reduced. As frequency is increased, the opposition to the flow of current in the center of the wire increases. Current in the center of the wire becomes smaller and most of the electron flow is on the wire surface. When the frequency applied is 100 megahertz or higher, the electron movement in the center is so small that the center of the wire could be removed without any noticeable effect on current. You should be able to see that the effective cross-sectional area decreases as the frequency increases. Since resistance is inversely proportional to the cross-sectional area, the resistance will increase as the frequency is increased. Also, since power loss increases as resistance increases, power losses increase with an increase in frequency because of skin effect.

Copper losses can be minimized and conductivity increased in an rf line by plating the line with silver. Since silver is a better conductor than copper, most of the current will flow through the silver layer. The tubing then serves primarily as a mechanical support.

Dielectric Losses

DIELECTRIC LOSSES result from the heating effect on the dielectric material between the conductors. Power from the source is used in heating the dielectric. The heat produced is dissipated into the surrounding medium. When there is no potential difference between two conductors, the atoms in the dielectric material between them are normal and the orbits of the electrons are circular. When there is a potential difference between two conductors, the orbits of the electrons change. The excessive negative charge on one conductor repels electrons on the dielectric toward the positive conductor and thus distorts the orbits of the electrons. A change in the path of electrons requires more energy, introducing a power loss.

The atomic structure of rubber is more difficult to distort than the structure of some other dielectric materials. The atoms of materials, such as polyethylene, distort easily. Therefore, polyethylene is often used as a dielectric because less power is consumed when its electron orbits are distorted.

Radiation and Induction Losses

RADIAION and INDUCTION LOSSES are similar in that both are caused by the fields surrounding the conductors. Induction losses occur when the electromagnetic field about a conductor cuts through any nearby metallic object and a current is induced in that object. As a result, power is dissipated in the object and is lost.

Radiation losses occur because some magnetic lines of force about a conductor do not return to the conductor when the cycle alternates. These lines of force are projected into space as radiation, and this results in power losses. That is, power is supplied by the source, but is not available to the load.

VOLTAGE CHANGE

In an electric circuit, energy is stored in electric and magnetic fields. These fields must be brought to the load to transmit that energy. At the load, energy contained in the fields is converted to the desired form of energy.

Transmission of Energy

When the load is connected directly to the source of energy, or when the transmission line is short, problems concerning current and voltage can be solved by applying Ohm's law. When the transmission line becomes long enough so the time difference between a change occurring at the generator and a change appearing at the load becomes appreciable, analysis of the transmission line becomes important.

Dc Applied to a Transmission Line

In figure 3-7, a battery is connected through a relatively long two-wire transmission line to a load at the far end of the line. At the instant the switch

is closed, neither current nor voltage exists on the line. When the switch is closed, point A becomes a positive potential, and point B becomes negative. These points of difference in potential move down the line. However, as the initial points of potential leave points A and B, they are followed by new points of difference in potential, which the battery adds at A and B. This is merely saying that the battery maintains a constant potential difference between points A and B. A short time after the switch is closed, the initial points of difference in potential have reached points A' and B'; the wire sections from points A to A' and points B to B' are at the same potential as A and B, respectively. The points of charge are represented by plus (+) and minus (-) signs along the wires, The directions of the currents in the wires are represented by the arrowheads on the line, and the direction of travel is indicated by an arrow below the line. Conventional lines of force represent the electric field that exists between the opposite kinds of charge on the wire sections from A to A' and B to B'. Crosses (tails of arrows) indicate the magnetic field created by the electric field moving down the line. The moving electric field and the accompanying magnetic field constitute an electromagnetic wave that is moving from the generator (battery) toward the load. This wave travels at approximately the speed of light in free space. The energy reaching the load is equal to that developed at the battery (assuming there are no losses in the transmission line). If the load absorbs all of the energy, the current and voltage will be evenly distributed along the line.

Ac Applied to a Transmission Line

When the battery of figure 3-7 is replaced by an ac generator (fig. 3-8), each successive instantaneous value of the generator voltage is propagated down the

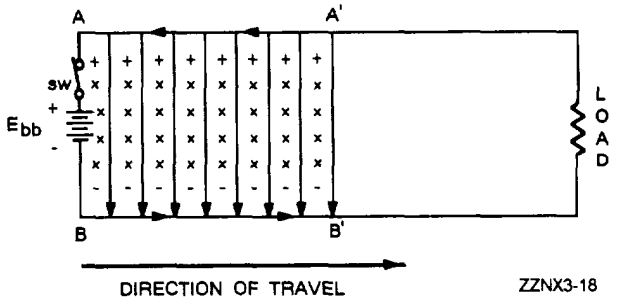


Figure 3-7.—Dc voltage applied to a line.

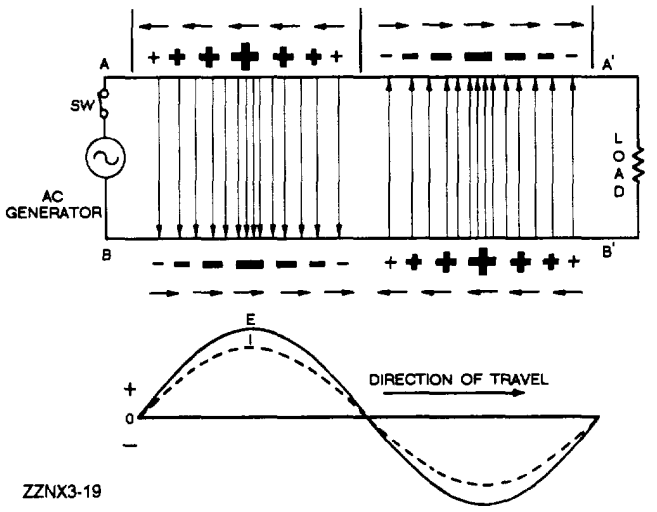


Figure 3-8.—Ac voltage applied to a line.

line at the speed of light. The action is similar to the wave created by the battery, except the applied voltage is sinusoidal instead of constant. Assume that the switch is closed at the moment the generator voltage is passing through zero and that the next half cycle makes point A positive. At the end of one cycle of generator voltage, the current and voltage distribution will be as shown in figure 3-8.

In this illustration the conventional lines of force represent the electric fields. For simplicity, the magnetic fields are not shown. Points of charge are indicated by plus (+) and minus (-) signs, the larger signs indicating points of higher amplitude of both voltage and current. Short arrows indicate direction of current (electron flow). The waveform drawn below the transmission line represents the voltage (E) and current (I) waves. The line is assumed to be infinite in length so there is no reflection. Thus, traveling sinusoidal voltage and current waves continually travel in phase from the generator toward the load, or far end of the line. Waves traveling from the generator to the load are called INCIDENT WAVES. Waves traveling from the load back to the generator are called REFLECTED WAVES and will be explained in later paragraphs.

STANDING-WAVE RATIO

The measurement of standing waves on a transmission line yields information about equipment operating

conditions. Maximum power is absorbed by the load when $Z_L = Z_0$. If a line has no standing waves, the termination for that line is correct and maximum power transfer takes place.

You have probably noticed that the variation of standing waves shows how near the rf line is to being terminated in Z_0. A wide variation in voltage along the length means a termination far from Z_0. A small variation means termination near Z_0. Therefore, the ratio of the maximum to the minimum is a measure of the perfection of the termination of a line. This ratio is called the STANDING-WAVE RATIO (swr) and is always expressed in whole numbers. For example, a ratio of 1:1 describes a line terminated in its characteristic impedance (Z_0).

Voltage Standing-Wave Ratio

The ratio of maximum voltage to minimum voltage on a line is called the VOLTAGE STANDING-WAVE RATIO (vswr). Therefore:

$$\text{vswr} = \left| \frac{E_{max}}{E_{min}} \right|$$

The vertical lines in the formula indicate that the enclosed quantities are absolute and that the two values are taken without regard to polarity, Depending on the nature of the standing waves, the numerical value of vswr ranges from a value of 1 ($Z_L = Z_0$, no standing waves) to an infinite value for theoretically complete reflection. Since there is always a small loss on a line, the minimum voltage is never zero and the vswr is always some finite value. However, if the vswr is to be a useful quantity. the power losses along the line must be small in comparison to the transmitted power.

Power Standing-Wave Ratio

The square of the vswr is called the POWER STANDING-WAVE RATIO (pswr). Therefore:

$$\text{pswr} = \frac{P_{max}}{P_{min}}$$

This ratio is useful because the instruments used to detect standing waves react to the square of the voltage. Since power is proportional to the square of the voltage, the ratio of the square of the maximum and minimum voltages is called the power standing-wave ratio. In a sense, the name is misleading because the power along a transmission line does not vary.

Current Standing-Wave Ratio

The ratio of maximum to minimum current along a transmission line is called CURRENT STANDING- WAVE RATIO (iswr). Therefore:

$$\text{iswr} = \left| \frac{I_{max}}{I_{min}} \right|$$

This ratio is the same as that for voltages. It can be used where measurements are made with loops that sample the magnetic field along a line. It gives the same results as vswr measurements.

TRANSMISSION MEDIUMS

The Navy uses many different types of TRANSMISSION MEDIUMS in its electronic applications. Each medium (line or waveguide) has a certain characteristic impedance value, current-carrying capacity, and physical shape and is designed to meet a particular requirement.

The five types of transmission mediums that we will discuss in this topic include PARALLEL-LINE, TWISTED PAIR, SHIELDED PAIR, COAXIAL LINE, and WAVEGUIDES. The use of a particular line depends, among other things, on the applied frequency, the power-handling capabilities, and the type of installation.

Parallel Line

One type of parallel line is the TWO-WIRE OPEN LINE, illustrated in figure 3-9. This line consists of two wires that are generally spaced from 2 to 6 inches apart by insulating spacers. This type of line is most often used for power lines, rural telephone lines, and telegraph lines. It is sometimes used as a transmission line between a transmitter and an antenna or between an antenna and a receiver. An advantage of this type

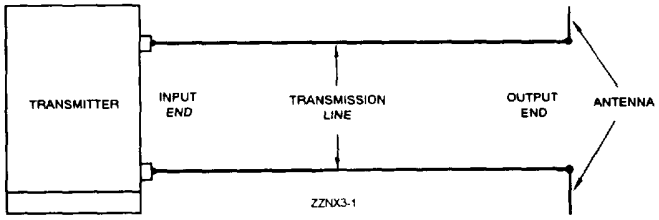


Figure 3-9.—Two-wire open line.

of line is its simple construction. The principal disadvantages of this type of line are the high radiation losses and electrical noise pickup because of the lack of shielding. Radiation losses are produced by the changing fields created by the changing current in each conductor.

Another type of parallel line is the TWO-WIRE RIBBON (TWIN LEAD) LINE, illustrated in figure 3-10. This type of transmission line is commonly used to connect a television receiving antenna to a home television set. This line is essentially the same as the two-wire open line except that uniform spacing is assured by embedding the two wires in a low-loss dielectric, usually polyethylene. Since the wires are embedded in the thin ribbon of polyethylene, the dielectric space is partly air and partly polyethylene.

Twisted Pair

The TWISTED PAIR transmission line is illustrated in figure 3-11. As the name implies, the line consists of two insulated wires twisted together to form a flexible line without the use of spacers. It is not used for transmitting high frequency because of the high dielectric losses that occur in the rubber insulation. When the line is wet, the losses increase greatly.

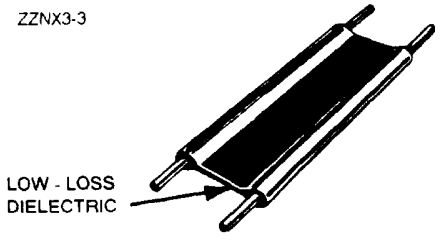


Figure 3-10.—Two-wire ribbon line.

Figure 3-11.—Twisted pair.

Shielded Pair

The SHIELDED PAIR, shown in figure 3-12, consists of parallel conductors separated from each other and surrounded by a solid dielectric. The conductors are contained within a braided copper tubing that acts as an electrical shield. The assembly is covered with a rubber or flexible composition coating that protects the line from moisture and mechanical damage. Outwardly, it looks much like the power cord of a washing machine or refrigerator.

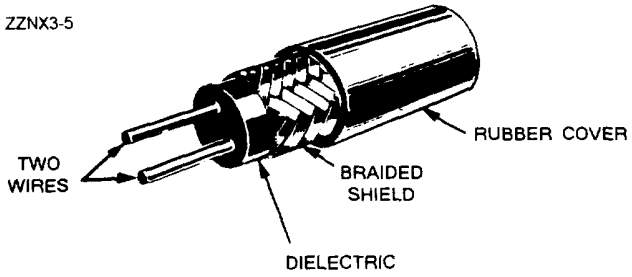


Figure 3-12.—Shielded pair.

The principal advantage of the shielded pair is that the conductors are balanced to ground; that is, the capacitance between the wires is uniform throughout the length of the line. This balance is due to the uniform spacing of the grounded shield that surrounds the wires along their entire length. The braided copper shield isolates the conductors from stray magnetic fields.

Coaxial Lines

There are two types of COAXIAL LINES, RIGID (AIR) COAXIAL LINE and FLEXIBLE (SOLID) COAXIAL LINE. The physical construction of both types is basically the same; that is, each contains two concentric conductors.

The rigid coaxial line consists of a central, insulated wire (inner conductor) mounted inside a tubular outer conductor. This line is shown in figure 3-13. In some applications, the inner conductor is also tubular. The inner conductor is insulated from the outer conductor by insulating spacers or beads at regular intervals. The spacers are made of Pyrex, polystyrene, or some other material that has good insulating characteristics and low dielectric losses at high frequencies.

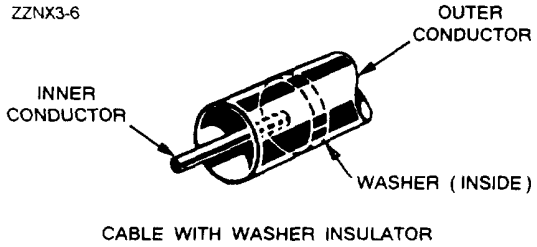


Figure 3-13.—Air coaxial line.

The chief advantage of the rigid line is its ability to minimize radiation losses. The electric and magnetic fields in a two-wire parallel line extend into space for relatively great distances and radiation losses occur. However, in a coaxial line no electric or magnetic fields extend outside of the outer conductor. The fields are confined to the space between the two conductors, resulting in a perfectly shielded coaxial line. Another advantage is that interference from other lines is reduced.

The rigid line has the following disadvantages: (1) it is expensive to construct; (2) it must be kept dry to prevent excessive leakage between the two conductors; and (3) although high-frequency losses are somewhat less than in previously mentioned lines, they are still excessive enough to limit the practical length of the line.

Leakage caused by the condensation of moisture is prevented in some rigid line applications by the use of an inert gas, such as nitrogen, helium, or argon. It is pumped into the dielectric space of the line at a pressure that can vary from 3 to 35 pounds per square inch. The inert gas is used to dry the line when it is first installed and pressure is maintained to ensure that no moisture enters the line.

Flexible coaxial lines (fig. 3-14) are made with an inner conductor that consists of flexible wire insulated from the outer conductor by a solid, continuous insulating material. The outer conductor is made of metal braid, which gives the line flexibility. Early attempts at gaining flexibility involved using rubber insulators between the two conductors. However, the rubber insulators caused excessive losses at high frequencies.

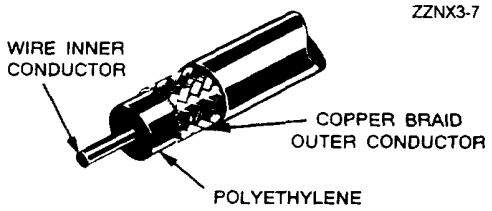


Figure 3-14.—Flexible coaxial line.

Because of the high-frequency losses associated with rubber insulators, polyethylene plastic was developed to replace rubber and eliminate these losses. Polyethylene plastic is a solid substance that remains flexible over a wide range of temperatures. It is unaffected by seawater, gasoline, oil, and most other liquids that may be found aboard ship. The use of polyethylene as an insulator results in greater high-frequency losses than the use of air as an insulator. However, these losses are still lower than the losses associated with most other solid dielectric materials.

This concludes our study of transmission lines. The rest of this chapter will be an introduction into the study of waveguides.

WAVEGUIDE THEORY

The two-wire transmission line used in conventional circuits is inefficient for transferring electromagnetic energy at microwave frequencies. At these frequencies, energy escapes by radiation because the fields are not confined in all directions, as illustrated in figure 3-15. Coaxial lines are more efficient than two-wire lines for transferring electromagnetic energy because the fields are completely confined by the conductors, as illustrated in figure 3-16. Waveguides are the most

efficient way to transfer electromagnetic energy. WAVEGUIDES are essentially coaxial lines without center conductors. They are constructed from conductive material and may be rectangular, circular, or elliptical in shape, as shown in figure 3-17.

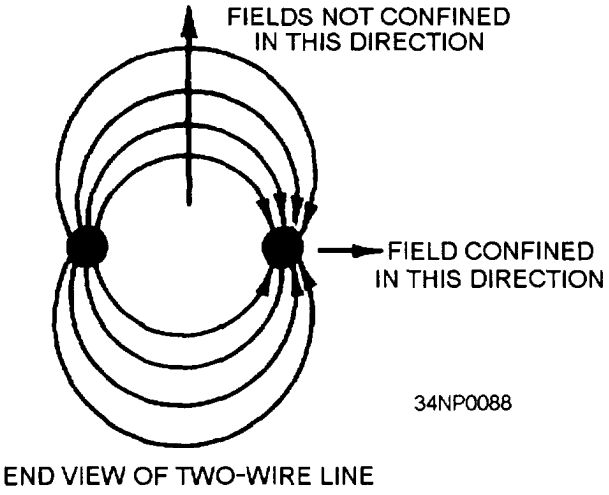


Figure 3-15.—Fields confined in two directions only.

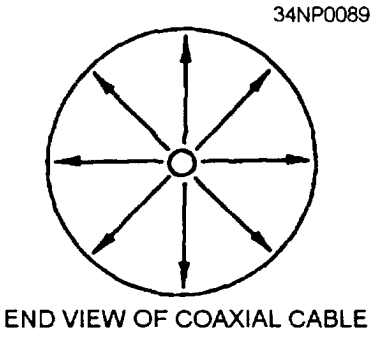


Figure 3-16.—Fields confined in all directions.

WAVEGUIDE ADVANTAGES

Waveguides have several advantages over two-wire and coaxial transmission lines. For example, the large surface area of waveguides greatly reduces COPPER (1^2R) LOSSES. Two-wire transmission lines have large copper losses because they have a relatively small surface area. The surface area of the outer conductor

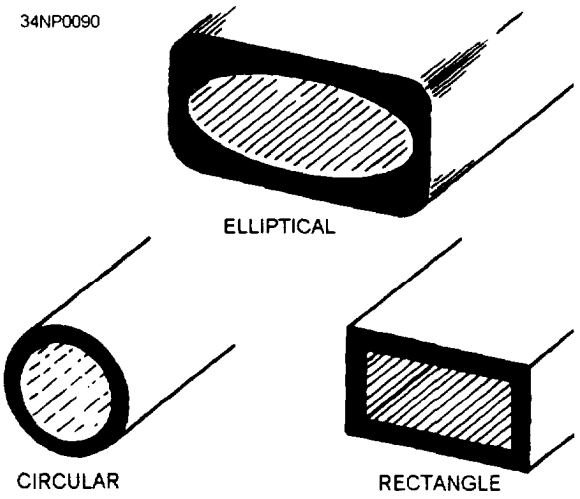


Figure 3-17.—Waveguide shapes.

of a coaxial cable is large, but the surface area of the inner conductor is relatively small. At microwave frequencies, the current-carrying area of the inner conductor is restricted to a very small layer at the surface of the conductor by an action called SKIN EFFECT.

Skin effect tends to increase the effective resistance of the conductor. Although energy transfer in coaxial cable is caused by electromagnetic field motion, the magnitude of the field is limited by the size of the current-carrying area of the inner conductor. The small size of the center conductor is even further reduced by skin effect, and energy transmission by coaxial cable becomes less efficient than by waveguides. DIELECTRIC LOSSES are also lower in waveguides than in two-wire and coaxial transmission lines. Dielectric losses in two-wire and coaxial lines are caused by the heating of the insulation between the conductors. The insulation behaves as the dielectric of a capacitor formed by the two wires of the transmission line. A voltage potential across the two wires causes heating of the dielectric and results in a power loss. In practical applications, the actual breakdown of the insulation between the conductors of a transmission line is more frequently a problem than is the dielectric loss.

This breakdown is usually caused by stationary voltage spikes or “nodes,” which are caused by standing waves. Standing waves are stationary and occur when part of the energy traveling down the line

is reflected by an impedance mismatch with the load. The voltage potential of the standing waves at the points of greatest magnitude can become large enough to break down the insulation between transmission line conductors.

The dielectric in waveguides is air, which has a much lower dielectric loss than conventional insulating materials. However, waveguides are also subject to dielectric breakdown caused by standing waves. Standing waves in waveguides cause arcing, which decreases the efficiency of energy transfer and can severely damage the waveguide. Also since the electromagnetic fields are completely contained within the waveguide, radiation losses are kept very low.

Power-handling capability is another advantage of waveguides. Waveguides can handle more power than coaxial lines of the same size because power-handling capability is directly related to the distance between conductors. Figure 3-18 illustrates the greater distance between conductors in a waveguide.

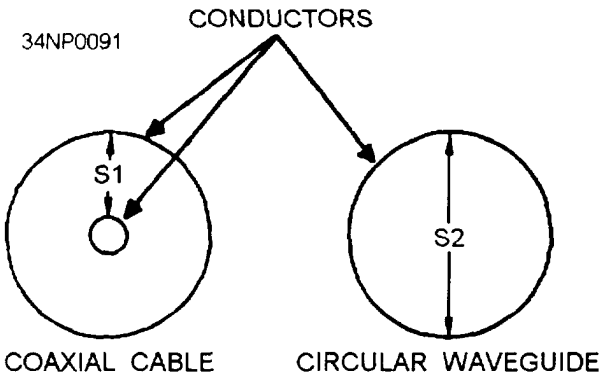


Figure 3-18.—Comparison of spacing in coaxial cable and a circular waveguide.

In view of the advantages of waveguides, you would think that waveguides should be the only type of transmission lines used. However, waveguides have certain disadvantages that make them practical for use only at microwave frequencies.

WAVEGUIDE DISADVANTAGES

Physical size is the primary lower-frequency limitation of waveguides. The width of a waveguide must be approximately a half wavelength at the frequency of the wave to be transported. For example, a waveguide for use at 1 megahertz would be about 700 feet wide. This makes the use of waveguides at frequencies below 1000 megahertz increasingly impractical. The lower frequency range of any system using waveguides is limited by the physical dimensions of the waveguides.

Waveguides are difficult to install because of their rigid, hollow-pipe shape. Special couplings at the joints are required to assure proper operation. Also, the inside surfaces of waveguides are often plated with silver or gold to reduce skin effect losses. These requirements increase the costs and decrease the practicality of waveguide systems at any other than microwave frequencies.

DEVELOPING THE WAVEGUIDE FROM PARALLEL LINES

You may better understand the transition from ordinary transmission line concepts to waveguide theories by considering the development of a waveguide from a two-wire transmission line. Figure 3-19 shows a section of a two-wire transmission line supported on two insulators. At the junction with the line, the insulators must present a very high impedance to ground for proper operation of the line. A low impedance insulator would obviously short-circuit the line to ground, and this is what happens at very high frequencies. Ordinary insulators display the characteristics of the dielectric of a capacitor formed by the wire and ground. As the frequency increases, the overall impedance decreases. A better high-frequency insulator is a quarter-wave section of transmission line shorted at one end. Such an insulator is shown in figure 3-20. The impedance of a shorted quarter-wave section is very high at the open-end junction with the two-wire transmission line. This type of insulator is known as a METALLIC INSULATOR and may be placed anywhere along a two-wire line.

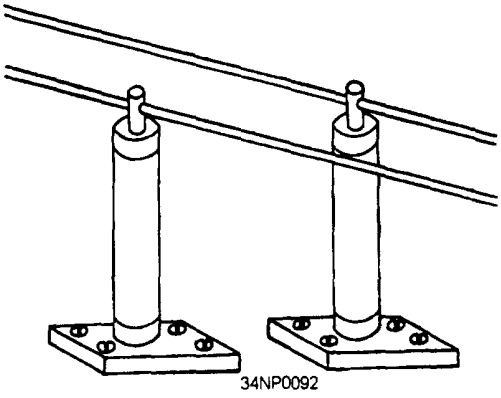


Figure 3-19.—Two-wire transmission line.

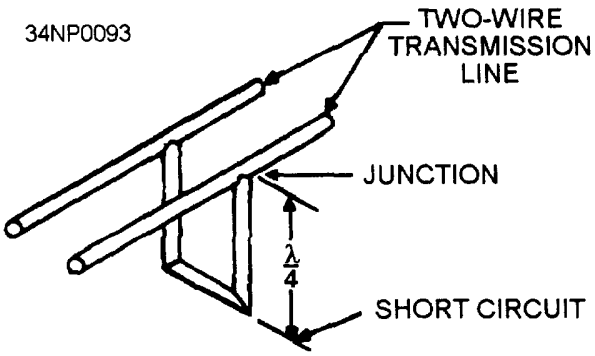


Figure 3-20.—Quarter-wave section of transmission line shorted at one end.

Note that quarter-wave sections are insulators at only one frequency. This severely limits the bandwidth, efficiency, and application of this type of two-wire line.

Figure 3-21 shows several metallic insulators on each side of a two-wire transmission line. As more insulators are added, each section makes contact with the next, and a rectangular waveguide is formed. The lines become part of the walls of the waveguide, as illustrated in figure 3-22. The energy is then conducted within the hollow waveguide instead of along the two-wire transmission line.

The comparison of the way electromagnetic fields work on a transmission line and in a waveguide is not exact. During the change from a two-wire line to a waveguide, the electromagnetic field configurations also undergo many changes. As a result of these changes, the waveguide does not actually operate

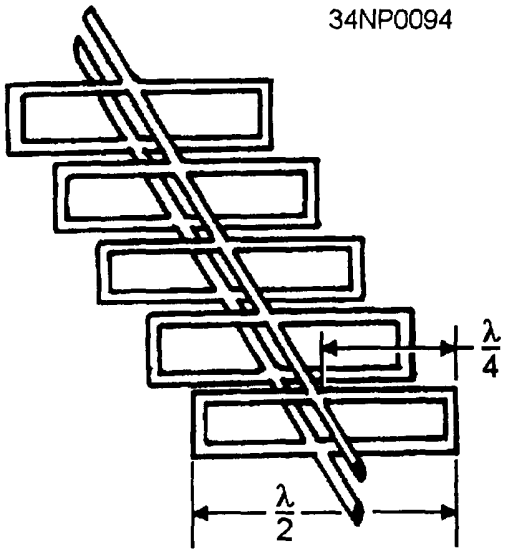


Figure 3-21.—Metallic insulator on each side of a two-wire line.

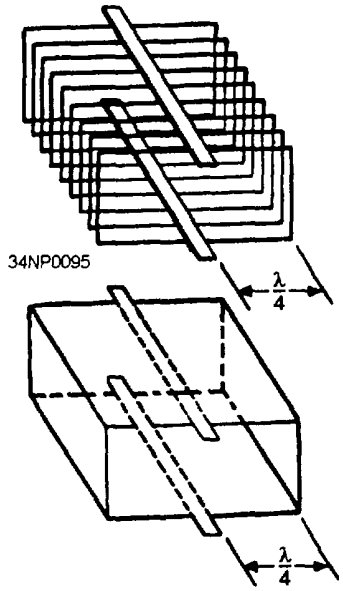


Figure 3-22.—Forming a waveguide by adding quarter-wave sections.

like a two-wire line that is completely shunted by quarter-wave sections. If it did, the use of a waveguide would be limited to a single-frequency wave length that was four times the length of the quarter-wave sections. In fact, waves of this length cannot pass efficiently through waveguides. Only a small range of frequencies of somewhat shorter wavelength (higher frequency) can pass efficiently.

As shown in figure 3-23, the widest dimension of a waveguide is called the “a” dimension and determines the range of operating frequencies. The narrowest dimension determines the power-handling capability of the waveguide and is called the “b” dimension.

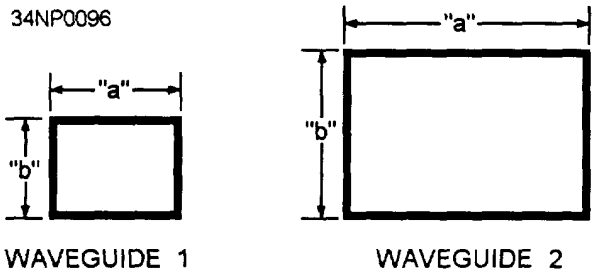


Figure 3-23.—Labeling waveguide dimensions,

NOTE: This method of labeling waveguides is not standard in all texts, Different methods may be used in other texts on microwave principles, but this method is in accordance with Navy Military Standards (MIL-STDS).

In theory, a waveguide could function at an infinite number of frequencies higher than the designed frequency; however, in practice, an upper frequency limit is caused by modes of operation, which will be discussed later.

If the frequency of a signal is decreased so much that two quarter-wavelengths are longer than the wide dimension of a waveguide, energy will no longer pass through the waveguide. This is the lower frequency limit, or CUTOFF FREQUENCY of a given waveguide. In practical applications, the wide dimension of a waveguide is usually 0.7 wavelength at the operating frequency. This allows the waveguide to handle a small range of frequencies both above and below the operating frequency. The “b” dimension is governed by the breakdown potential of the dielectric, which is usually air. Dimensions ranging from 0.2 to 0.5 wavelength are common for the “b” sides of a waveguide.

ENERGY PROPAGATION IN WAVEGUIDES

Since energy is transferred through waveguides by electromagnetic fields, you need a basic understanding of field theory. Both electric (E FIELD) and magnetic fields (H FIELD) are present in waveguides, and the interaction of these fields causes energy to travel through the waveguide. This action is best understood by first looking at the properties of the two individual fields.

E Field

An electric field exists when a difference of potential causes a stress in the dielectric between two points. The simplest electric field is one that forms between the plates of a capacitor when one plate is made positive compared to the other, as shown in view A of figure 3-24. The stress created in the dielectric is an electric field.

Electric fields are represented by arrows that point from the positive toward the negative potential. The number of arrows shows the relative strength of the field. In view B, for example, evenly spaced arrows indicate the field is evenly distributed. For ease of explanation, the electric field is abbreviated E field, and the lines of stress are called E lines.

H Field

The magnetic field in a waveguide is made up of magnetic lines of force that are caused by current flow through the conductive material of the waveguide. Magnetic lines of force, called H lines, are continuous closed loops, as shown in figure 3-25. All of the H lines associated with current are collectively called a magnetic field or H field. The strength of the H field, indicated by the number of H lines in a given area, varies directly with the amount of current.

Although H lines encircle a single, straight wire, they behave differently when the wire is formed into a coil, as shown in figure 3-26. In a coil the individual H lines tend to form around each turn of wire. Since

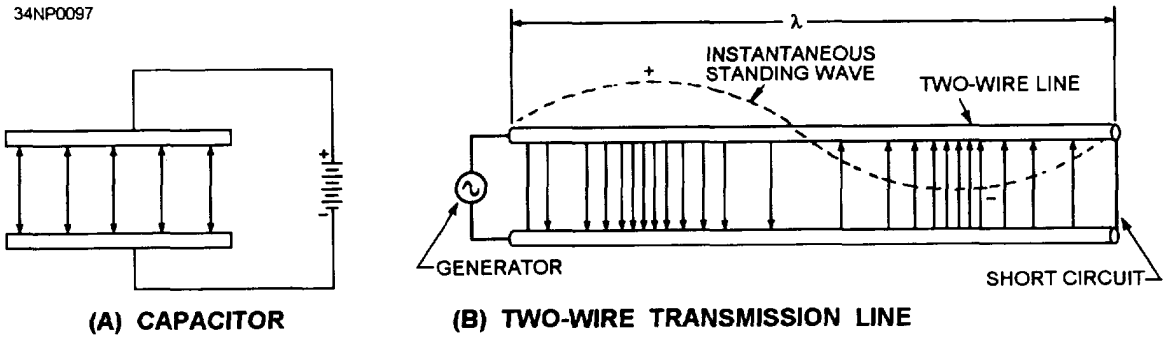


Figure 3-24.—Simple electric fields.

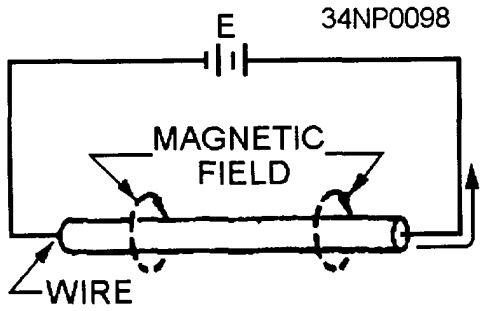


Figure 3-25.—Magnetic field on a single wire.

the H lines take opposite directions between adjacent turns, the field between the turns is canceled. Inside and outside the coil, where the direction of each H field is the same, the fields join and form continuous H lines around the entire coil. A similar action takes place in a waveguide.

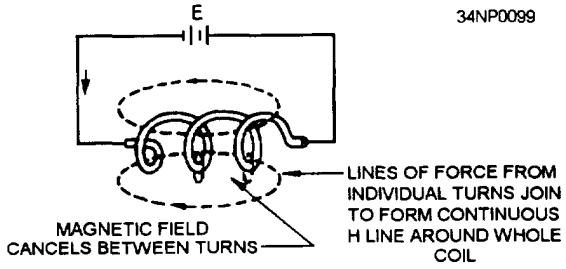


Figure 3-26.—Magnetic field on a coil.

BOUNDARY CONDITIONS IN A WAVEGUIDE

The travel of energy down a waveguide is similar, but not identical, to the travel of electromagnetic waves in free space. The difference is that the energy in a waveguide is confined to the physical limits of the guide. Two conditions, known as BOUNDARY CONDITIONS, must be satisfied for energy to travel through a waveguide.

The first boundary condition (illustrated in fig. 3-27, view A can be stated as follows:

For an electric field to exist at the surface of a conductor, it must be perpendicular to the conductor.

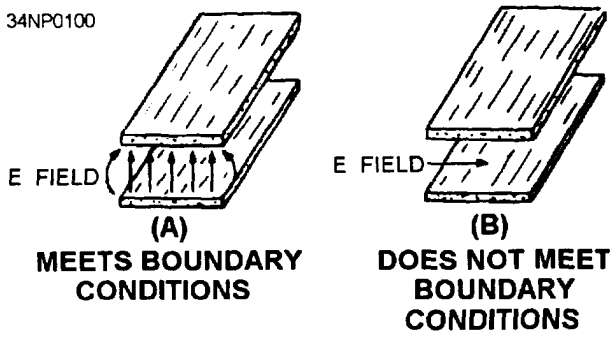


Figure 3-27.—E field boundary condition.

The opposite of this boundary condition, shown in view B, is also true. An electric field CANNOT exist parallel to a perfect conductor.

The second boundary condition, which is illustrated in figure 3-28, can be stated as follows:

For a varying magnetic field to exist, it must form closed loops in parallel with the conductors and be perpendicular to the electric field.

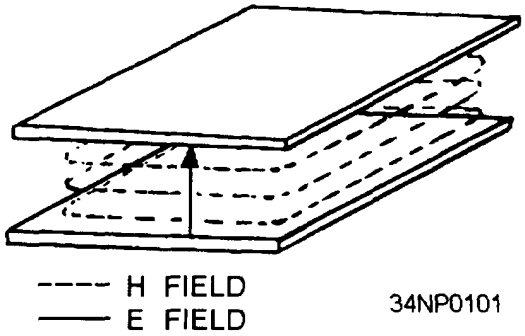


Figure 3-28.—H field boundary condition.

Since an E field causes a current flow that in turn produces an H field, both fields always exist at the same time in a waveguide. If a system satisfies one of these boundary conditions, it must also satisfy the other since neither field can exist alone.

WAVEFRONTS WITHIN A WAVEGUIDE

Electromagnetic energy transmitted into space consists of electric and magnetic fields that are at right angles (90 degrees) to each other and at right angles to the direction of propagation. A simple analogy to establish this relationship is by use of the right-hand rule for electromagnetic energy, based on the POYNTING VECTOR. It indicates that a screw (right-hand thread) with its axis perpendicular to the electric and magnetic fields will advance in the direction of propagation if the E field is rotated to the right (toward the H field). This rule is illustrated in figure 3-29.

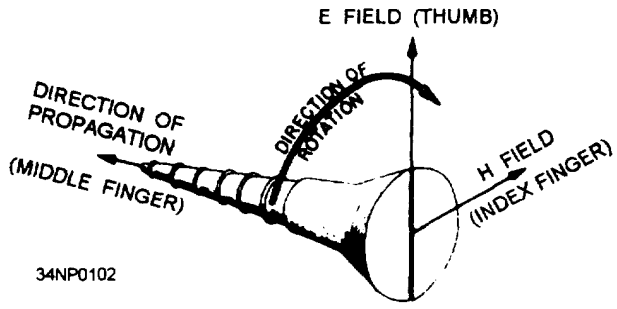


Figure 3-29.—The Poynting vector.

The combined electric and magnetic fields form a wavefront that can be represented by alternate negative and positive peaks at half-wavelength intervals, as illustrated in figure 3-30. Angle $\varnothing$ is the direction of travel of the wave with respect to some reference axis.

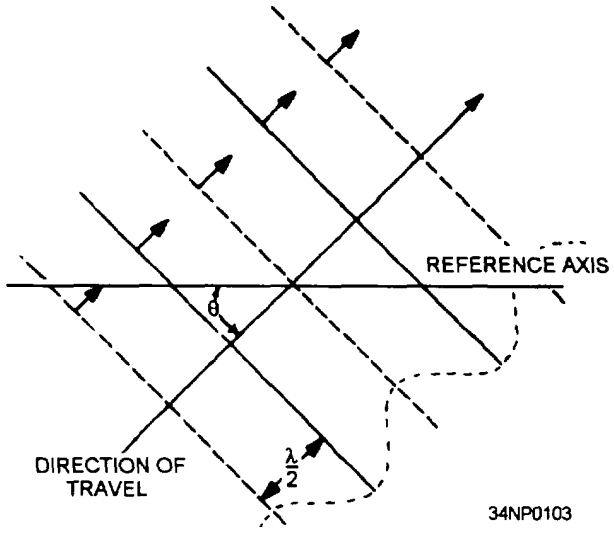


Figure 3-30.—Wavefronts in space.

The reflection of a single wavefront off the "b" wall of a waveguide is shown in figure 3-31. The wavefront is shown in view A as small particles, In views B and C particle 1 strikes the wall and is bounced back from the wall without losing velocity. If the wall is perfectly flat, the angle at which it the wall, known as the angle of incidence (θ), is the same as the angle of reflection $(\varnothing)$. An instant after particle 1 strikes the wall, particle 2 strikes the wall, as shown

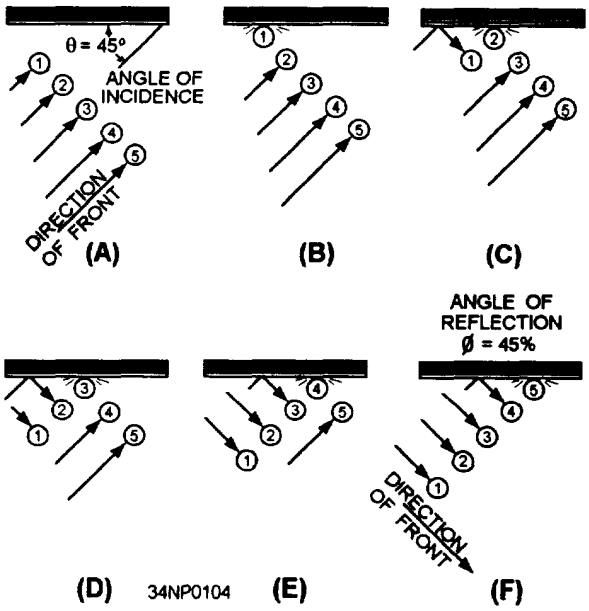


Figure 3-31.—Reflection of a single wavefront.

in view C, and reflects in the same manner. Because all the particles are traveling at the same velocity, particles 1 and 2 do not change their relative position with respect to each other. Therefore, the reflected wave has the same shape as the original. The remaining particles as shown in views D, E, and F reflect in the same manner. This process results in a reflected wavefront identical in shape, but opposite in polarity, to the incident wave.

Figure 3-32, views A and B, each illustrate the direction of propagation of two different electromagnetic wavefronts of different frequencies being radiated into a waveguide by a probe. Note that only the direction of propagation is indicated by the lines and arrowheads. The wavefronts are at right angles to the direction of propagation. The angle of incidence (θ) and the angle of reflection $(\varnothing)$ of the wavefronts vary in size with the frequency of the input energy, but the angles of reflection are equal to each other in a waveguide. The CUTOFF FREQUENCY in a waveguide is a frequency that would cause angles of incidence and reflection to be perpendicular to the walls of the guide. At any frequency below the cutoff frequency, the wavefronts will be reflected back and forth across the guide (setting up standing waves) and no energy will be conducted down the waveguide.

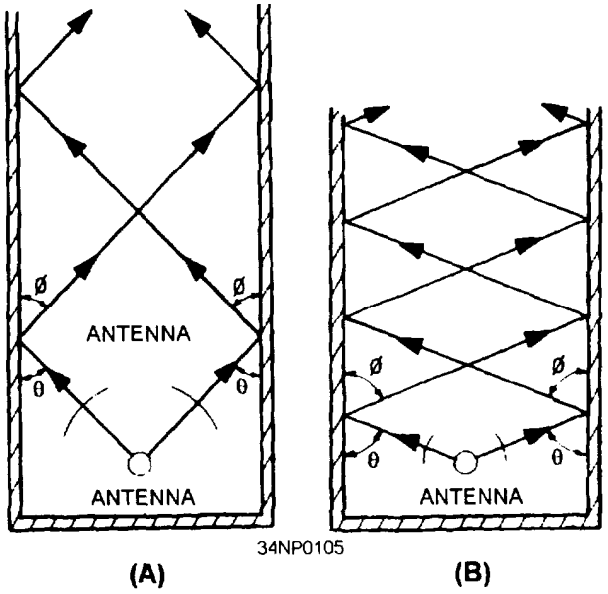


Figure 3-32.—Different frequencies in a waveguide.

The velocity of propagation of a wave along a waveguide is less than its velocity through free space (speed of light). This lower velocity is caused by the zigzag path taken by the wavefront. The forward-progress velocity of the wavefront in a waveguide is called GROUP VELOCITY and is somewhat slower than the speed of light.

The group velocity of energy in a waveguide is determined by the reflection angle of the wavefronts off the "b" walls. The reflection angle is determined by the frequency of the input energy. This basic principle is illustrated in figure 3-33. As frequency is decreased. the reflection angle increases, causing the group velocity to decrease. The opposite is also true; increasing frequency increases the group velocity.

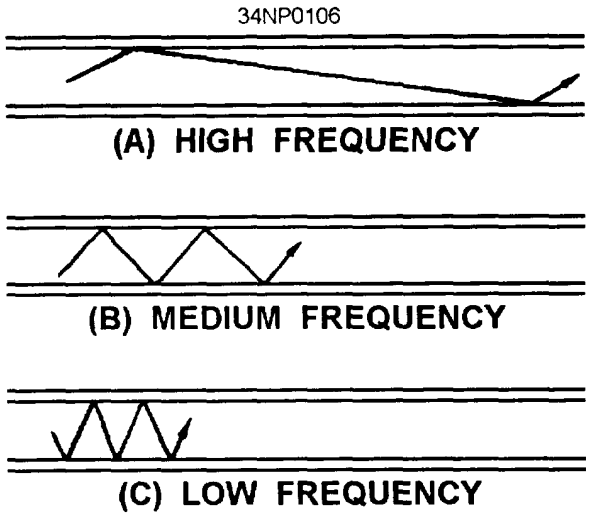


Figure 3-33.—Reflection angle at various frequencies.

WAVEGUIDE MODES OF OPERATION

The waveguide analyzed in the previous paragraphs yields an electric field configuration known as the half-sine electric distribution. This configuration, called a MODE OF OPERATION, is shown in figure 3-34. Recall that the strength of the field is indicated by the spacing of the lines; that is, the closer the lines, the stronger the field. The regions of maximum voltage in this field move continuously down the waveguide in a sine-wave pattern. To meet boundary conditions. the field must always be zero at the "b" walls.

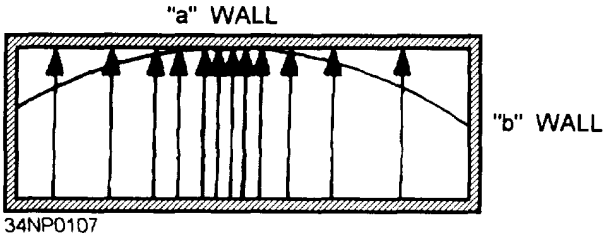


Figure 3-34.—Half-sine E field distribution.

The half-sine field is only one of many field configurations, or modes, that can exist in a rectangular waveguide. A full-sine field can also exist in a rectangular waveguide because, as shown in figure 3-35, the field is zero at the “b” walls.

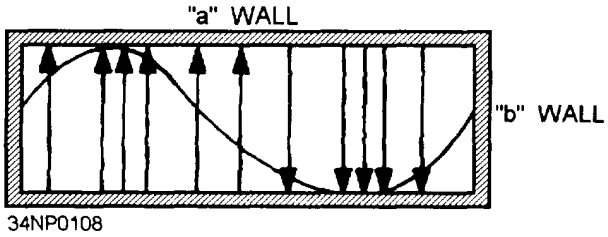


Figure 3-35.—Full-sine E field distribution.

The magnetic field in a rectangular waveguide is in the form of closed loops parallel to the surface of the conductors. The strength of the magnetic field is proportional to the electric field. Figure 3-36 illustrates the magnetic field pattern associated with a half-sine electric field distribution. The magnitude of the magnetic field varies in a sine-wave pattern down the center of the waveguide in “time phase” with the electric field. TIME PHASE means that the peak H lines and peak E lines occur at the same instant in time, although not necessarily at the same point along the length of the waveguide.

The dominant mode is the most efficient mode. Waveguides are normally designed so that only the dominant mode will be used. To operate in the dominant mode, a waveguide must have an “a” (wide) dimension of at least one half-wavelength of the frequency to be propagated. The “a” dimension of the waveguide must be kept near the minimum allowable value to ensure that only the dominant mode will exist. In practice, this dimension is usually 0.7 wavelength.

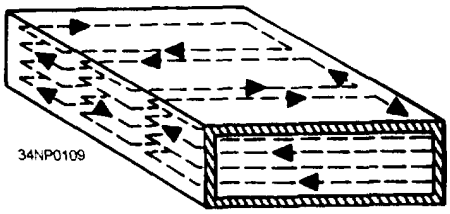


Figure 3-36.—Magnetic field caused by a half-sine E field.

Of the possible modes of operation available for a given waveguide, the dominant mode has the lowest cutoff frequency. The high-frequency limit of a rectangular waveguide is a frequency at which its “a” dimension becomes large enough to allow operation in a mode higher than that for which the waveguide has been designed.

Circular waveguides are used in specific areas of radar and communications systems, such as rotating joints used at the mechanical point where the antennas rotate. Figure 3-37 illustrates the dominant mode of a circular waveguide. The cutoff wavelength of a circular guide is 1.71 times the diameter of the waveguide. Since the “a” dimension of a rectangular waveguide is approximately one half-wavelength at the cutoff frequency, the diameter of an equivalent circular waveguide must be 2/1.71, or approximately

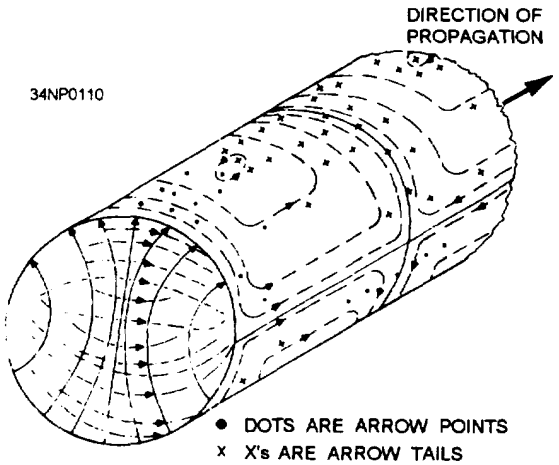


Figure 3-37.—Dominant mode in a circular waveguide.

1.17 times the "a" dimension of a rectangular waveguide.

MODE NUMBERING SYSTEMS

So far, only the most basic types of E and H field arrangements have been shown. More complicated arrangements are often necessary to make possible coupling, isolation, or other types of operation. The field arrangements of the various modes of operation are divided into two categories: TRANSVERSE ELECTRIC (TE) and TRANSVERSE MAGNETIC (TM).

In the transverse electric (TE) mode, the entire electric field is in the transverse plane, which is perpendicular to the waveguide, (direction of energy travel). Part of the magnetic field is parallel to the length axis.

In the transverse magnetic (TM) mode, the entire magnetic field is in the transverse plane and has no portion parallel to the length axis.

Since there are several TE and TM modes, subscripts are used to complete the description of the field pattern. In rectangular waveguides, the first subscript indicates the number of half-wave patterns in the "a" dimension, and the second subscript indicates the number of half-wave patterns in the "b" dimension.

The dominant mode for rectangular waveguides is shown in figure 3-38. It is designated as the TE mode because the E fields are perpendicular to the "a" walls. The first subscript is 1, since there is only one half-wave pattern across the "a" dimension. There are no E-field patterns across the "b" dimension, so the second subscript is 0. The complete mode description of the dominant mode in rectangular waveguides is $TE_{1,0}$. Subsequent description of waveguide operation in this text will assume the dominant ($TE_{1,0}$) mode unless otherwise noted.

A similar system is used to identify the modes of circular waveguides. The general classification of TE and TM is true for both circular and rectangular waveguides. In circular waveguides the subscripts have a different meaning. The first subscript indicates the number of fill-wave patterns around the circumference of the waveguide. The second subscript indicates the number of half-wave patterns across the diameter.

In the circular waveguide in figure 3-39, the E field is perpendicular to the length of the waveguide with no E lines parallel to the direction of propagation. Thus, it must be classified as operating in the TE mode. If you follow the E line pattern in a counterclockwise direction starting at the top, the E lines go from zero, through maximum positive (tail of arrows), back to zero, through maximum negative (head of arrows), and then back to zero again. This is one full wave, so the first subscript is 1. Along the diameter, the E lines go from zero through maximum and back to zero, making a half-wave variation. The second subscript, therefore, is also 1. $TE_{1,1}$ is the complete mode description of the dominant mode in circular waveguides. Several modes are possible in both circular and rectangular waveguides. Figure 3-40 illustrates several different modes that can be used to verify the mode numbering system.

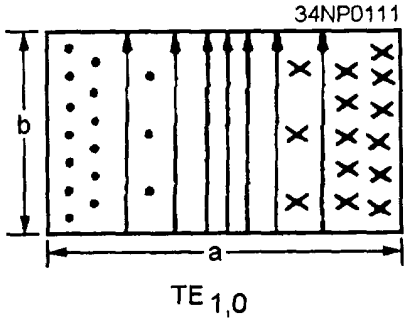


Figure 3-38.—Dominant mode in a rectangular waveguide.

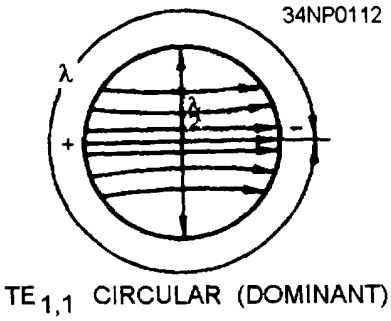


Figure 3-39.—Counting wavelengths in a circular waveguide.

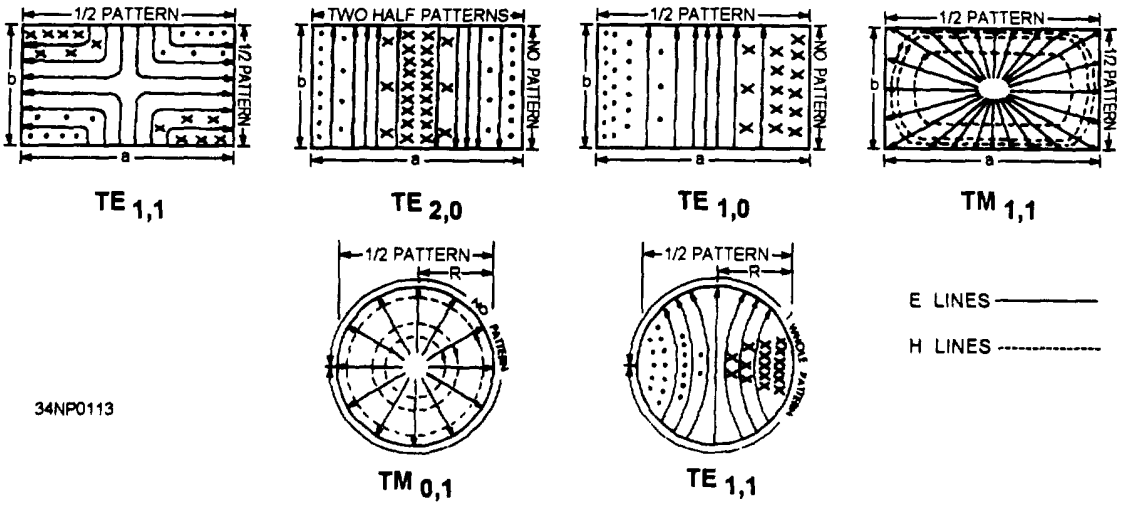


Figure 3-40.—Various modes of operation for rectangular and circular waveguides.

WAVEGUIDE INPUT/OUTPUT METHODS

A waveguide, as explained earlier in this topic, operates differently from an ordinary transmission line. Therefore, special devices must be used to put energy into a waveguide at one end and remove it from the other end.

The three devices used to injector remove energy from waveguides are PROBES, LOOPS, and SLOTS. Slots may also be called APERTURES or WINDOWS.

When a small probe is inserted into a waveguide and supplied with microwave energy, it acts as a quarter-wave antenna. Current flows in the probe and sets up an E field such as the one shown in figure 3-41, view A. The E lines detach themselves from the probe. When the probe is located at the point of highest efficiency, the E lines set up an E field of considerable intensity.

The most efficient place to locate the probe is in the center of the “a” wall, parallel to the “b” wall, and one quarter-wavelength from the shorted end of the waveguide, as shown in figure 3-41, views B and C. This is the point at which the E field is maximum in the dominant mode. Therefore, energy transfer (coupling) is maximum at this point. Note that the quarter-wavelength spacing is at the frequency required to propagate the dominant mode.

In many applications a lesser degree of energy transfer, called loose coupling, is desirable. The amount of energy transfer can be reduced by decreasing the length of the probe, by moving it out of the center of the E field, or by shielding it. Where the degree of coupling must be varied frequently, the probe is made retractable so the length can be easily changed.

The size and shape of the probe determines its frequency, bandwidth, and power-handling capability. As the diameter of a probe increases, the bandwidth increases. A probe similar in shape to a door knob is capable of handling much higher power and a larger bandwidth than a conventional probe. The greater power-handling capability is directly related to the increased surface area. Two examples of broad-bandwidth probes are illustrated in figure 3-41, view D. Removal of energy from a waveguide is simply a reversal of the injection process using the same type of probe.

Another way of injecting energy into a waveguide is by setting up an H field in the waveguide. This can be accomplished by inserting a small loop that carries a high current into the waveguide, as shown in figure 3-42, view A. A magnetic field builds up around the loop and expands to fit the waveguide, as shown in view B. If the frequency of the current in the loop is within the bandwidth of the waveguide, energy will be transferred to the waveguide.

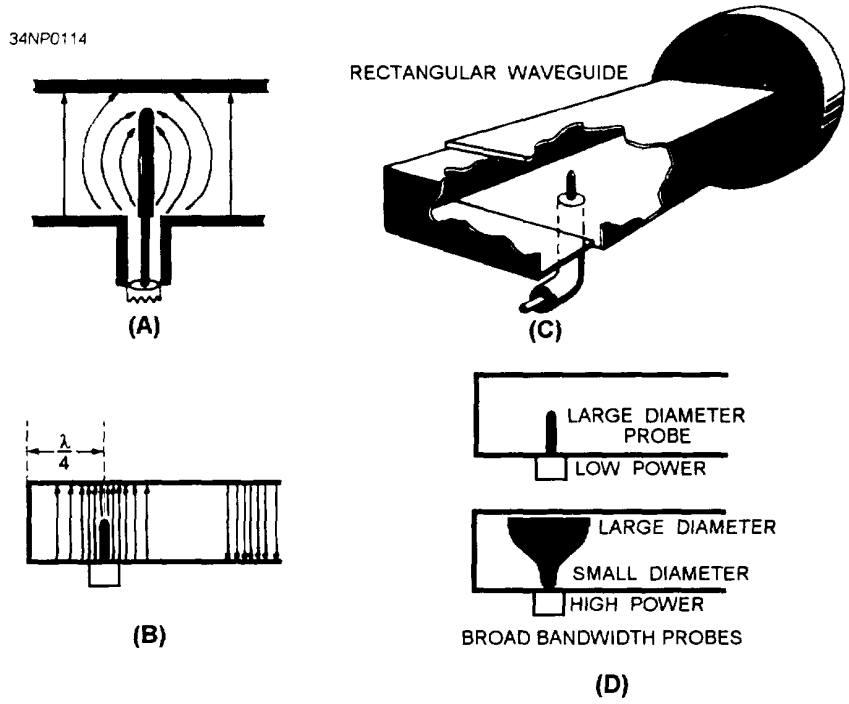


Figure 3-41.—Probe coupling in a rectangular waveguide.

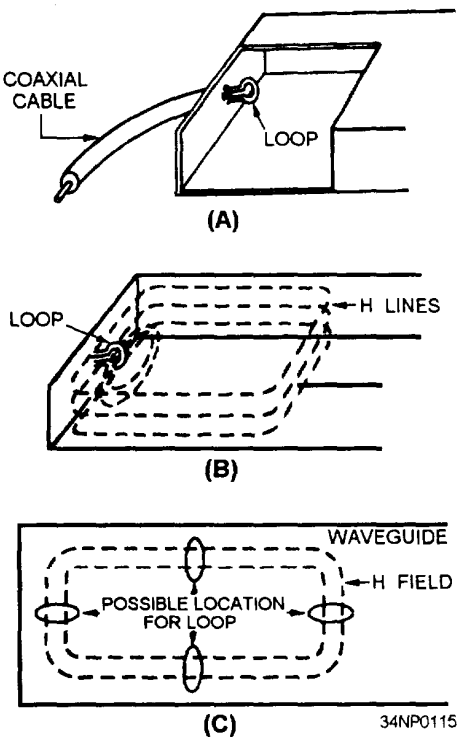


Figure 3-42.—Loop coupling in a rectangular waveguide.

For the most efficient coupling to the waveguide, the loop is inserted at one of several points where the magnetic field will be of greatest strength. Four of those points are shown in figure 3-42, view C.

When less efficient coupling is desired, you can rotate or move the loop until it encircles a smaller number of H lines. When the diameter of the loop is increased, its power-handling capability also increases. The bandwidth can be increased by increasing the size of the wire used to make the loop.

When a loop is introduced into a waveguide in which an H field is present, a current is induced in the loop. When this condition exists, energy is removed from the waveguide.

Slots or apertures are sometimes used when very loose (inefficient) coupling is desired, as shown in figure 3-43. In this method energy enters through a small slot in the waveguide and the E field expands into the waveguide. The E lines expand first across the slot and then across the interior of the waveguide.

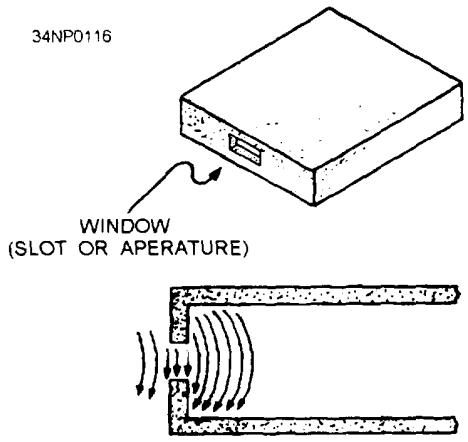


Figure 3-43.—Slot coupling in a waveguide.

Minimum reflections occur when energy is injected or removed if the size of the slot is properly proportioned to the frequency of the energy.

After learning how energy is coupled into and out of a waveguide with slots, you might think that leaving the end open is the most simple way of injecting or removing energy in a waveguide. This is not the case, however, because when energy leaves a waveguide, fields form around the end of the waveguide. These fields cause an impedance mismatch which, in turn, causes the development of standing waves and a drastic loss in efficiency. Various methods of impedance matching and terminating waveguides will be covered in the next section.

WAVEGUIDE IMPEDANCE MATCHING

Waveguide transmission systems are not always perfectly impedance matched to their load devices. The standing waves that result from a mismatch cause a power loss, a reduction in power-handling capability, and an increase in frequency sensitivity. Impedance-changing devices are therefore placed in the waveguide to match the waveguide to the load. These devices are placed near the source of the standing waves.

Figure 3-44 illustrates three devices, called irises, that are used to introduce inductance or capacitance into a waveguide. An iris is nothing more than a metal plate that contains an opening through which the waves may pass. The iris is located in the transverse plane of either the magnetic or electric field.

An inductive iris and its equivalent circuit are illustrated in figure 3-44, view A. The iris places a shunt inductive reactance across the waveguide that is directly proportional to the size of the opening. Notice that the inductive iris is in the magnetic plane. The shunt capacitive reactance, illustrated in view B, basically acts the same way. Again, the reactance is directly proportional to the size of the opening, but the iris is placed in the electric plane. The iris, illustrated in view C, has portions in both the magnetic

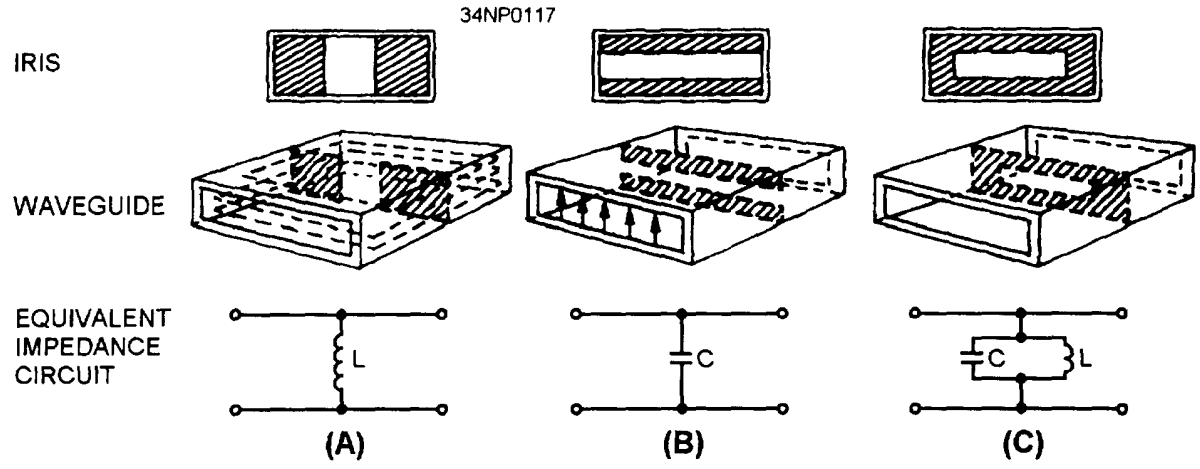


Figure 3-44.—Waveguide irises.

and electric transverse planes and forms an equivalent parallel-LC circuit across the waveguide. At the resonant frequency, the iris acts as a high shunt resistance. Above or below resonance, the iris acts as a capacitive or inductive reactance.

POSTS and SCREWS made from conductive material can be used for impedance-changing devices in waveguides. Views A and B of figure 3-45, illustrate two basic methods of using posts and screws. A post or screw that only partially penetrates into the waveguide acts as a shunt capacitive reactance. When the post or screw extends completely through the waveguide, making contact with the top and bottom walls, it acts as an inductive reactance. Note that when screws are used, the amount of reactance can be varied.

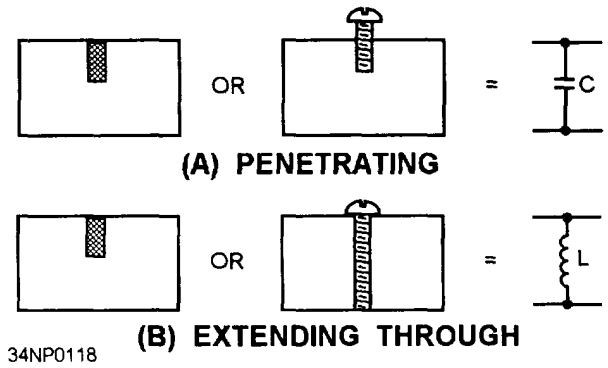


Figure 3-45.—Conducting posts and screws.

WAVEGUIDE TERMINATIONS

Electromagnetic energy is often passed through a waveguide to transfer the energy from a source into space. As previously mentioned, the impedance of a waveguide does not match the impedance of space, and without proper impedance matching standing waves cause a large decrease in the efficiency of the waveguide.

Any abrupt change in impedance causes standing waves, but when the change in impedance at the end of a waveguide is gradual, almost no standing waves are formed. Gradual changes in impedance can be obtained by terminating the waveguide with a funnel-shaped HORN, such as the three types illustrated in figure 3-46. The type of horn used depends upon the frequency and the desired radiation pattern.

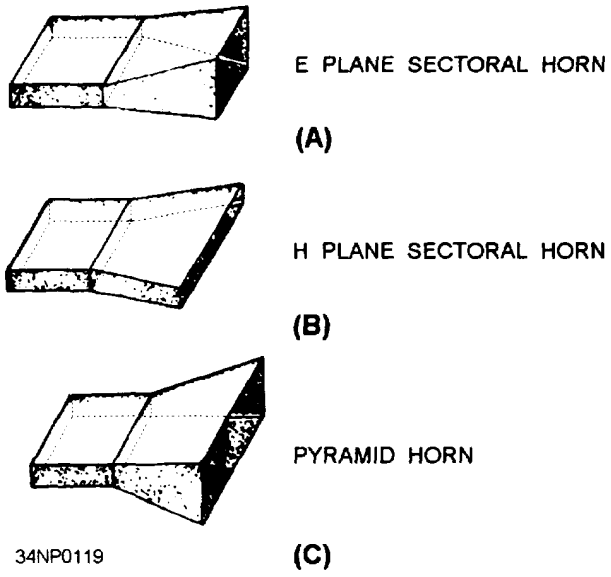


Figure 3-46.—Waveguide horns.

As you may have noticed, horns are really simple antennas. They have several advantages over other impedance-matching devices, such as their large bandwidth and simple construction.

A waveguide may also be terminated in a resistive load that is matched to the characteristic impedance of the waveguide. The resistive load is most often called a DUMMY LOAD, because its only purpose is to absorb all the energy in a waveguide without causing standing waves.

There is no place on a waveguide to connect a fixed termination resistor; therefore, several special arrangements are used to terminate waveguides. One method is to fill the end of the waveguide with a graphite and sand mixture, as illustrated in figure 3-47, view A. When the fields enter the mixture, they induce a current flow in the mixture that dissipates the energy as heat. Another method (view B) is to use a high-resistance rod placed at the center of the E field. The E field causes current to flow in the rod, and the high resistance of the rod dissipates the energy as a power loss, again in the form of heat.

Still another method for terminating a waveguide is the use of a wedge of highly resistive material, as shown in view C of figure 3-47. The plane of the wedge is placed perpendicular to the magnetic lines

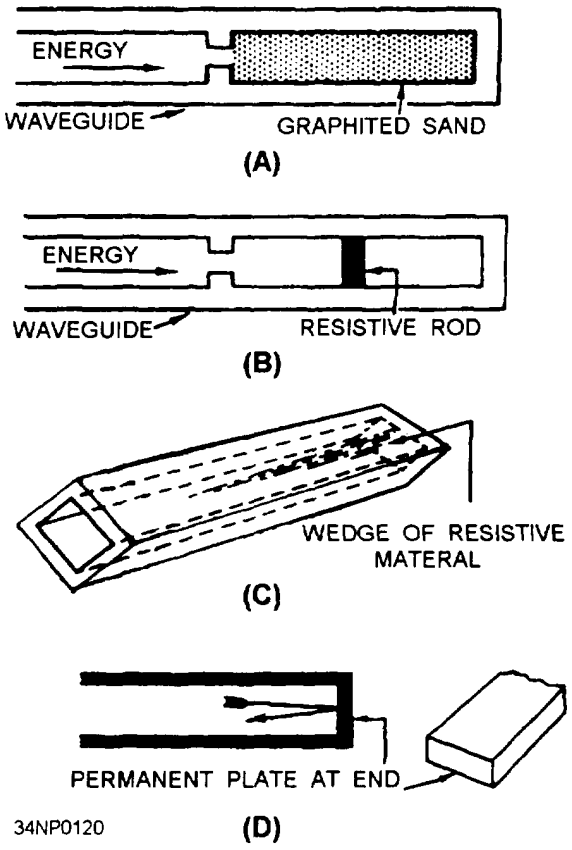


Figure 3-47.—Terminating waveguides.

of force. When the H lines cut through the wedge, current flows in the wedge and causes a power loss. As with the other methods, this loss is in the form of heat. Since very little energy reaches the end of the waveguide, reflections are minimum.

All of the terminations discussed so far are designed to radiate or absorb the energy without reflections. In many instances, however, all of the energy must be reflected from the end of the waveguide. The best way to accomplish this is to permanently weld a metal plate at the end of the waveguide, as shown in view D of figure 3-47.

WAVEGUIDE PLUMBING

Since waveguides are really only hollow metal pipes, the installation and the physical handling of waveguides have many similarities to ordinary plumbing. In light of this fact, the bending, twisting, joining, and installation of waveguides is commonly called waveguide plumbing. Naturally, waveguides are different in design from pipes that are designed to carry liquids or other substances. The design of a waveguide is determined by the frequency and power level of the electromagnetic energy it will carry. The following paragraphs explain the physical factors involved in the design of waveguides.

Waveguide Bends

The size, shape, and dielectric material of a waveguide must be constant throughout its length for energy to move from one end to the other without reflections. Any abrupt change in its size or shape can cause reflections and a loss in overall efficiency. When such a change is necessary, the bends, twists, and joints of the waveguides must meet certain conditions to prevent reflections.

Waveguides maybe bent in several ways that do not cause reflections. One way is the gradual bend shown in figure 3-48. This gradual bend is known as an E bend because it distorts the E fields. The E bend must have a radius greater than two wavelengths to prevent reflections.

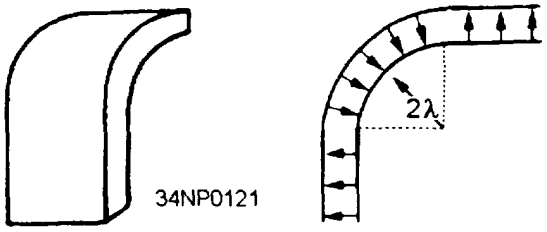


Figure 3-48.—Gradual E bend.

Another common bend is the gradual H bend (fig. 3-49). It is called an H bend because the H fields are distorted when a waveguide is bent in this manner. Again, the radius of the bend must be greater than two wavelengths to prevent reflections. Neither the E bend in the “a” dimension nor the H bend in the “b” dimension changes the normal mode of operation.

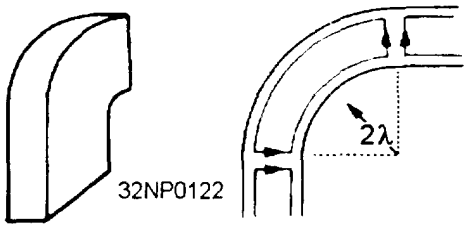


Figure 3-49.—Gradual H bend.

A sharp bend in either dimension may be used if it meets certain requirements. Notice the two 45-degree bends in figure 3-50; the bends are 1/4λ apart. The reflections that occur at the 45-degree bends cancel each other, leaving the fields as though no reflections have occurred.

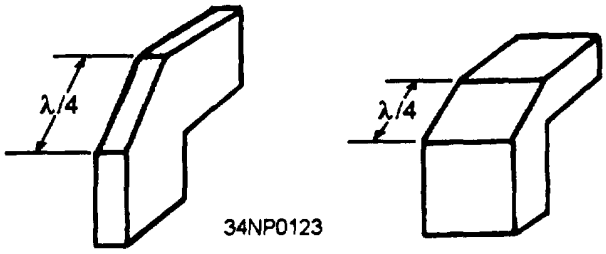


Figure 3-50.—Sharp bends.

Sometimes the electromagnetic fields must be rotated so that they are in the proper phase to match the phase of the load. This may be accomplished by twisting the waveguide as shown in figure 3-51. The twist must be gradual and greater than 2λ.

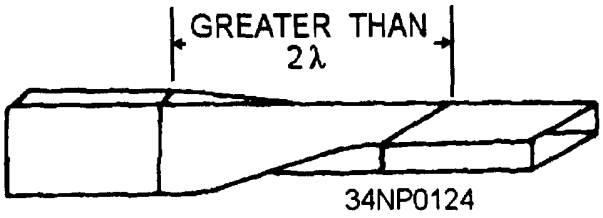


Figure 3-51.—Waveguide twist.

The flexible waveguide (fig. 3-52) allows special bends, which some equipment applications might require. It consists of a specially wound ribbon of conductive material, the most commonly used is brass, with the inner surface plated with chromium. Power losses are greater in the flexible waveguide because the inner surfaces are not perfectly smooth. Therefore, it is only used in short sections where no other reasonable solution is available.

Waveguide Joints

Since an entire waveguide system cannot possibly be molded into one piece, the waveguide must be

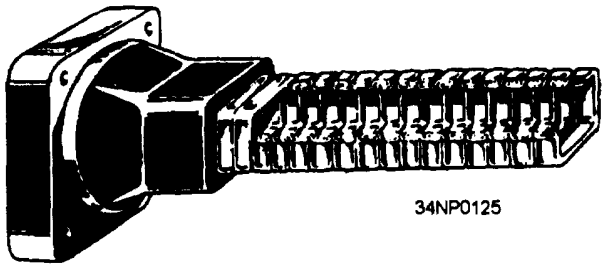


Figure 3-52.—Flexible waveguide.

constructed in sections and the sections connected with joints. The three basic types of waveguide joints are the PERMANENT, the SEMIPERMANENT, and the ROTATING JOINTS. Since the permanent joint is a factory-welded joint that requires no maintenance, only the semipermanent and rotating joints will be discussed.

Sections of waveguide must be taken apart for maintenance and repair. A semipermanent joint, called a CHOKE JOINT, is most commonly used for this purpose. The choke joint provides good electromagnetic continuity between the sections of the waveguide with very little power loss.

A cross-sectional view of a choke joint is shown in figure 3-53. The pressure gasket shown between the two metal surfaces forms an airtight seal. Notice in view B that the slot is exactly 1/4λ from the "a" wall of the waveguide. The slot is also 1/4λ deep, as shown in view A, and because it is shorted at point 1, a high impedance results at point 2. Point 3 is 1/4λ from point 2. The high impedance at point 2 results in a low impedance, or short, at point 3. This effect creates a good electrical connection between the two sections that permits energy to pass with very little reflection or loss.

Whenever a stationary rectangular waveguide is to be connected to a rotating antenna, a rotating joint must be used. A circular waveguide is normally used in a rotating joint. Rotating a rectangular waveguide would cause field pattern distortion. The rotating section of the joint, illustrated in figure 3-54, uses a choke joint to complete the electrical connection with the stationary section. The circular waveguide is designed so that it will operate in the $TM_{0,1}$ mode.

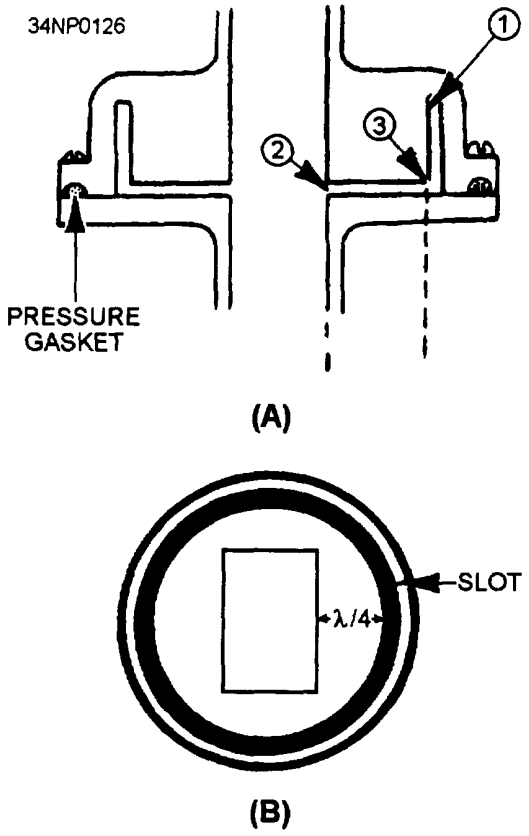


Figure 3-53.—Choke joint.

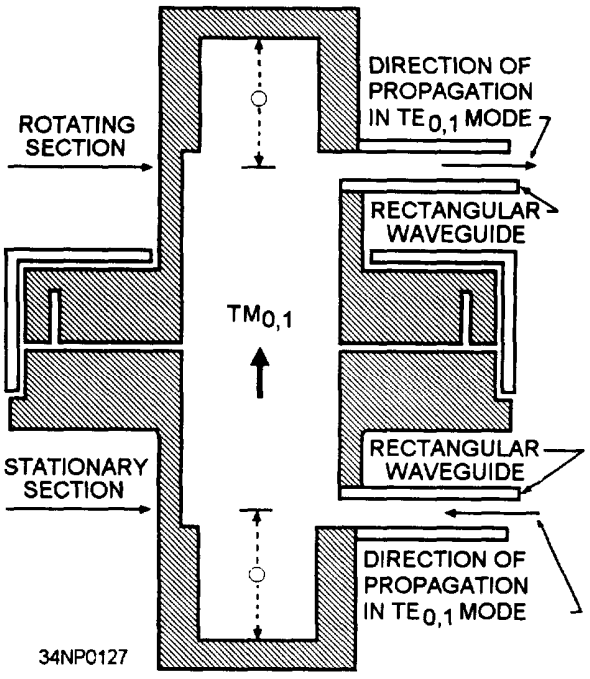


Figure 3-54.—Rotating joint.

The rectangular sections are attached as shown in the illustration to prevent the circular waveguide from operating in the wrong mode. Distance "O" is $3/4\lambda$ so that a high impedance will be presented to any unwanted modes. This is the most common design used for rotating joints, but other types may be used in specific applications.

WAVEGUIDE MAINTENANCE

The installation of a waveguide system presents problems that are not normally encountered when dealing with other types of transmission lines. These problems often fall within the technician's area of responsibility. A brief discussion of waveguide handling, installation, and maintenance will help prepare you for this maintenance responsibility, Detailed information concerning waveguide maintenance in a particular system may be found in the technical manuals for the system.

Since a waveguide naturally has a low loss ratio, most losses in a waveguide system are caused by other factors. Improperly connected joints or damaged inner surfaces can decrease the efficiency of a system to the point that it will not work at all. Therefore, you must take great care when working with waveguides to prevent physical damage. Since waveguides are made from a soft, conductive material, such as copper or aluminum, they are very easy to dent or deform. Even the slightest damage to the inner surface of a waveguide will cause standing waves and, often, internal arcing. Internal arcing causes further damage to the waveguide in an action that is often self-sustaining until the waveguide is damaged beyond use. Part of your job as a technician will be to inspect the waveguide system for physical damage. The previously mentioned dents are only one type of physical damage that can decrease the efficiency of the system. Another problem occurs because waveguides are made from a conductive material such as copper while the structures of most ships are made from steel. When two dissimilar metals, such as copper and steel, are in direct contact, an electrical action called ELECTROLYSIS takes place that causes very rapid corrosion of the metals. Waveguides can be completely destroyed by electrolytic corrosion in a relatively short period of time if they are not isolated from direct contact with other metals. Any inspection

of a waveguide system should include a detailed inspection of all support points to ensure that electrolytic corrosion is not taking place. Any waveguide that is exposed to the weather should be painted and all joints sealed. Proper painting prevents natural corrosion, and sealing the joints prevents moisture from entering the waveguide.

Moisture can be one of the worst enemies of a waveguide system. As previously discussed, the dielectric in waveguides is air, which is an excellent dielectric as long as it is free of moisture. Wet air, however, is a very poor dielectric and can cause serious internal arcing in a waveguide system. For this reason, care is taken to ensure that waveguide systems are pressurized with air that is dry. Checking the pressure and moisture content of the waveguide air may be one of your daily system maintenance duties.

More detailed waveguide installation and maintenance information can be found in the technical manuals that apply to your particular system. Another good source is the *Electronics Installation and Maintenance Handbooks (EIMB)* published by Naval Sea Systems Command. *Installation Standards (EIMB) Handbook,* NAVSEA 0967-LP-000-0110, is the volume that deals with waveguide installation and maintenance.

WAVEGUIDE DEVICES

The discussion of waveguides, up to this point, has been concerned only with the transfer of energy from one point to another. Many waveguide devices have been developed, however, that modify the energy in some fashion during the transmission. Some devices do nothing more than change the direction of the energy. Others have been designed to change the basic characteristics or power level of the electromagnetic energy.

This section will explain the basic operating principles of some of the more common waveguide devices, such as DIRECTIONAL COUPLERS, CAVITY RESONATORS, and HYBRID JUNCTIONS.

Directional Couplers

The directional coupler is a device that provides a method of sampling energy from within a waveguide for measurement or use in another circuit. Most couplers sample energy traveling in one direction only. However, directional couplers can be constructed that sample energy in both directions. These are called BIDIRECTIONAL couplers and are widely used in radar and communications systems.

Directional couplers may be constructed in many ways. The coupler illustrated in figure 3-55 is constructed from an enclosed waveguide section of the same dimensions as the waveguide in which the energy is to be sampled. The "b" wall of this enclosed section is mounted to the "b" wall of the waveguide from which the sample will be taken. There are two holes in the "b" wall between the sections of the coupler. These two holes are $1/4\lambda$ apart. The upper section of the directional coupler has a wedge of energy-absorbing material at one end and a pickup probe connected to an output jack at the other end. The absorbent material absorbs the energy not directed at the probe and a portion of the overall energy that enters the section.

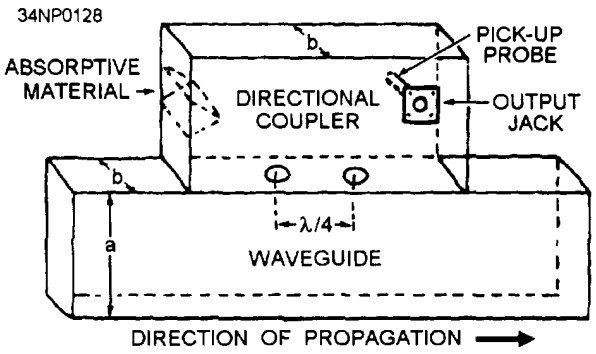


Figure 3-55.—Directional coupler.

Figure 3-56 illustrates two portions of the incident wavefront in a waveguide. The waves travel down the waveguide in the direction indicated and enter the coupler section through both holes. Since both portions of the wave travel the same distance, they are in phase when they arrive at the pickup probe. Because the waves are in phase, they add together and provide a sample of the energy traveling down the waveguide. The sample taken is only a small portion of the energy that is traveling down the waveguide. The magnitude of the sample, however, is proportional to the magnitude of the energy in the waveguide. The absorbent material is designed to ensure that the ratio

between the sample energy and the energy in the waveguide is constant. Otherwise, the sample would contain no useful information. The ratio is usually stamped on the coupler in the form of an attenuation factor.

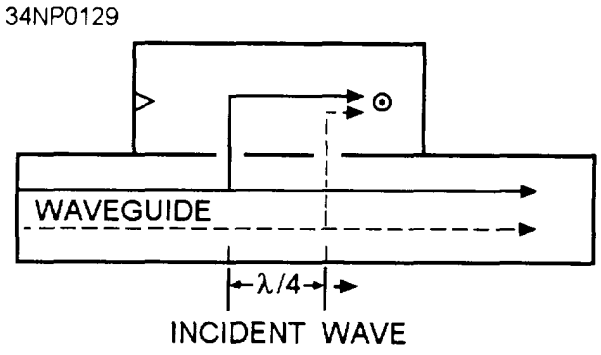


Figure 3-56.—Incident wave in a directional coupler designed to sample incident waves.

The effect of a directional coupler on any reflected energy is illustrated in figure 3-57. Note that these two waves do not travel the same distance to the pickup probe. The wave represented by the dotted line travels $1/2\lambda$ further and arrives at the probe 180 degrees out of phase with the wave, represented by the solid line. Because the waves are 180 degrees out of phase at the probe, they cancel each other and no energy is induced into the pickup probe. When the reflected energy arrives at the absorbent material, it adds and is absorbed by the material.

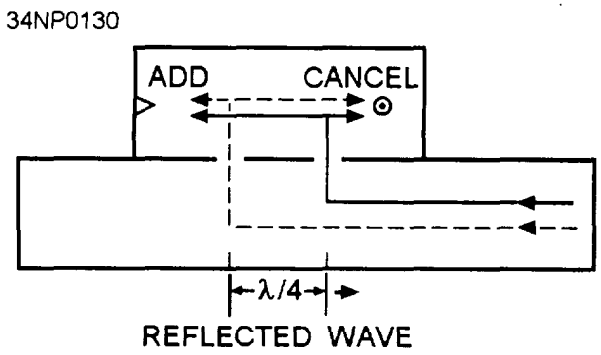


Figure 3-57.—Reflected wave in a directional coupler.

A directional coupler designed to sample reflected energy is shown in figure 3-58. The absorbent material and the probe are in opposite positions from the directional coupler designed to sample the incident energy. This positioning causes the two portions of the reflected energy to arrive at the probe in phase, providing a sample of the reflected energy. The transmitted energy is absorbed by the absorbent material.

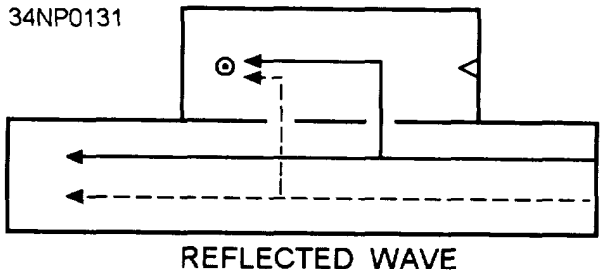


Figure 3-58.—Directional coupler designed to sample retlected energy.

A simple bidirectional coupler for sampling both transmitted and reflected energy can be constructed by mounting two directional couplers on opposite sides of a waveguide, as shown in figure 3-59.

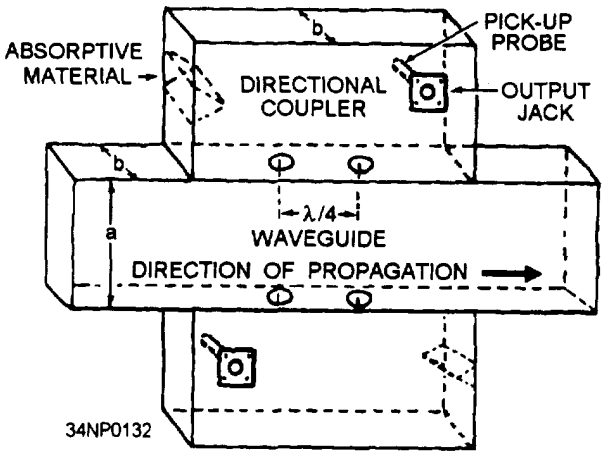


Figure 3-59.—Bidirectional coupler.

Cavity Resonators

By definition, a resonant cavity is any space completely enclosed by conducting walls that can contain oscillating electromagnetic fields and possess resonant properties. The cavity has many advantages

and uses at microwave frequencies. Resonant cavities have a very high Q and can be built to handle relatively large amounts of power. Cavities with a Q value in excess of 30,000 are not uncommon. The high Q gives these devices a narrow bandpass and allows very accurate tuning. Simple, rugged construction is an additional advantage.

Although cavity resonators, built for different frequency ranges and applications, have a variety of shapes, the basic principles of operation are the same for all.

One example of a cavity resonator is the rectangular box shown in figure 3-60, view A. It may be thought of as a section of rectangular waveguide closed at both ends by conducting plates. The frequency at which the resonant mode occurs is $1/2\lambda$ of the distance between the end plates. The magnetic field patterns in the rectangular cavity are shown in view B.

There are two variables that determine the primary frequency of any resonant cavity. The first variable is PHYSICAL SIZE. In general, the smaller the cavity, the higher its resonant frequency. The second controlling factor is the SHAPE of the cavity. Figure 3-61 illustrates several cavity shapes that are commonly used. Remember from the previously stated definition of a resonant cavity that any completely enclosed conductive surface, regardless of its shape, can act as a cavity resonator.

Energy can be inserted or removed from a cavity by the same methods that are used to couple energy into and out of waveguides. The operating principles of probes, loops, and slots are the same whether used in a cavity or a waveguide. Therefore, any of the three methods can be used with cavities to inject or remove energy.

The resonant frequency of a cavity can be varied by changing any of the three parameters: cavity volume, cavity capacitance, or cavity inductance. Changing the frequencies of a cavity is known as TUNING. The mechanical methods of tuning a cavity may vary with the application, but all methods use the same electrical principles.

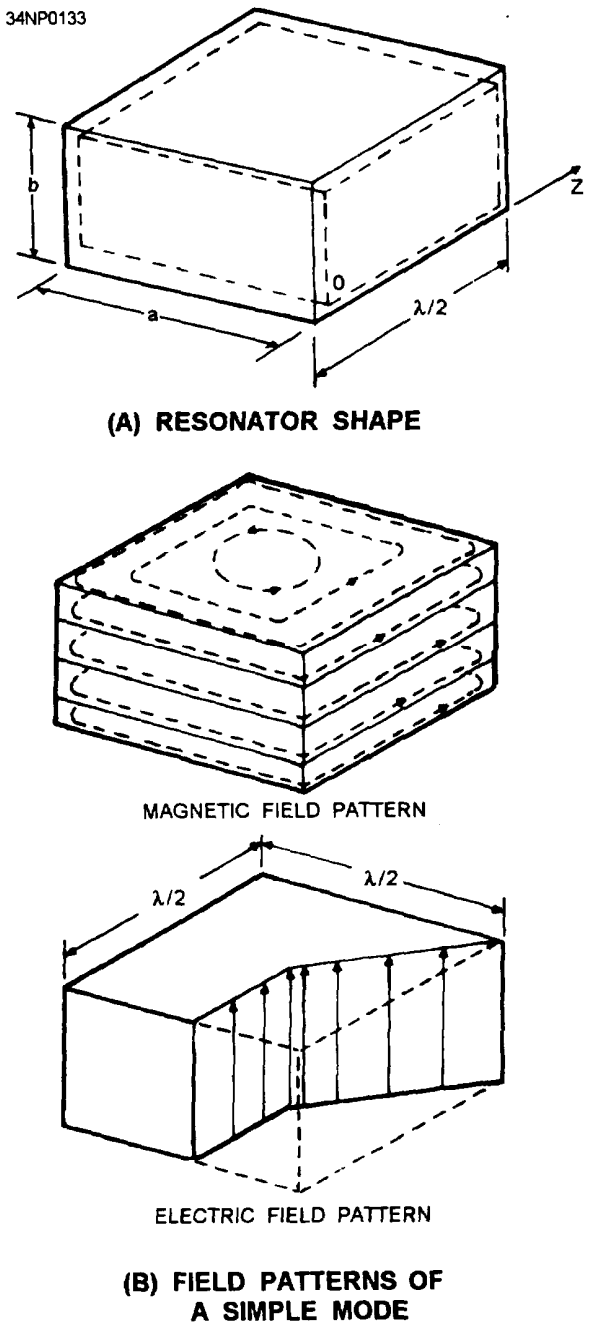


Figure 3-60.—Rectangular waveguide cavity resonator.

Waveguide Junctions

You may have assumed that when energy traveling down a waveguide reaches a junction it simply divides and follows the junction. This is not strictly true.

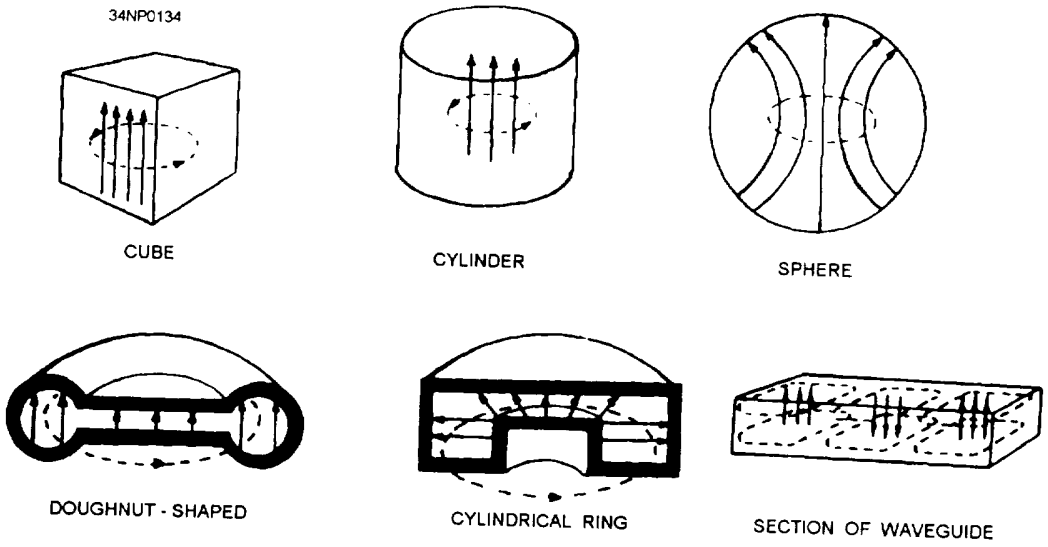


Figure 3-61.—Types of cavities.

Different types of junctions affect the energy in different ways. Since waveguide junctions are used extensively in most systems, you need to understand the basic operating principles of those most commonly used.

The T JUNCTION is the most simple of the commonly used waveguide junctions. T junctions are divided into two basic types, the E TYPE and the H TYPE. HYBRID JUNCTIONS are more complicated developments of the basic T junctions. The MAGIC-T and the HYBRID RING are the two most commonly used hybrid junctions.

E-TYPE T JUNCTION.— An E-type T junction is illustrated in figure 3-62, view A.

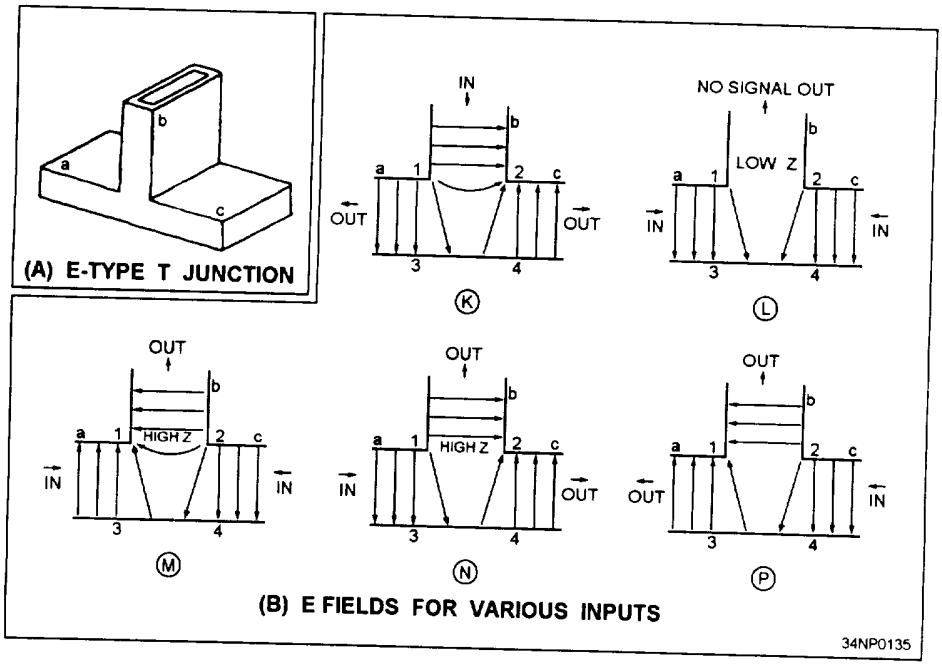


Figure 3-62.—E fields in an E-type T junction.

It is called an E-type T junction because the junction arm extends from the main waveguide in the same direction as the E field in the waveguide.

Figure 3-62, view B, illustrates cross-sectional views of the E-type T junction with inputs fed into the various arms. For simplicity, the magnetic lines that are always present with an electric field have been omitted. In view K, the input is fed into arm b and the outputs are taken from the a and c arms. When the E field arrives between points 1 and 2, point 1 becomes positive and point 2 becomes negative. The positive charge at point 1 then induces a negative charge on the wall at point 3. The negative charge at point 2 induces a positive charge at point 4. These charges cause the fields to form 180 degrees out of phase in the main waveguide; therefore, the outputs will be 180 degrees out of phase with each other. In view L, two in-phase inputs of equal amplitude are fed into the a and c arms. The signals at points 1 and 2 have the same phase and amplitude. No difference of potential exists across the entrance to the b arm, and no energy will be coupled out. However, when the two signals fed into the a and c arms are 180 degrees out of phase, as shown in view M, points 1 and 2 have a difference of potential. This difference of potential induces an E field from point 1 to point 2 in the b arm, and energy is coupled out of this arm. Views N and P illustrate two methods of obtaining two outputs with only one input.

H-TYPE T JUNCTION.— An H-type T junction is illustrated in figure 3-63, view A. It is called an H-type T junction because the long axis of the "b" arm is parallel to the plane of the magnetic lines of force in the waveguide. Again, for simplicity, only the E lines are shown in this figure. Each X indicates an E line moving away from the observer. Each dot indicates an E line moving toward the observer.

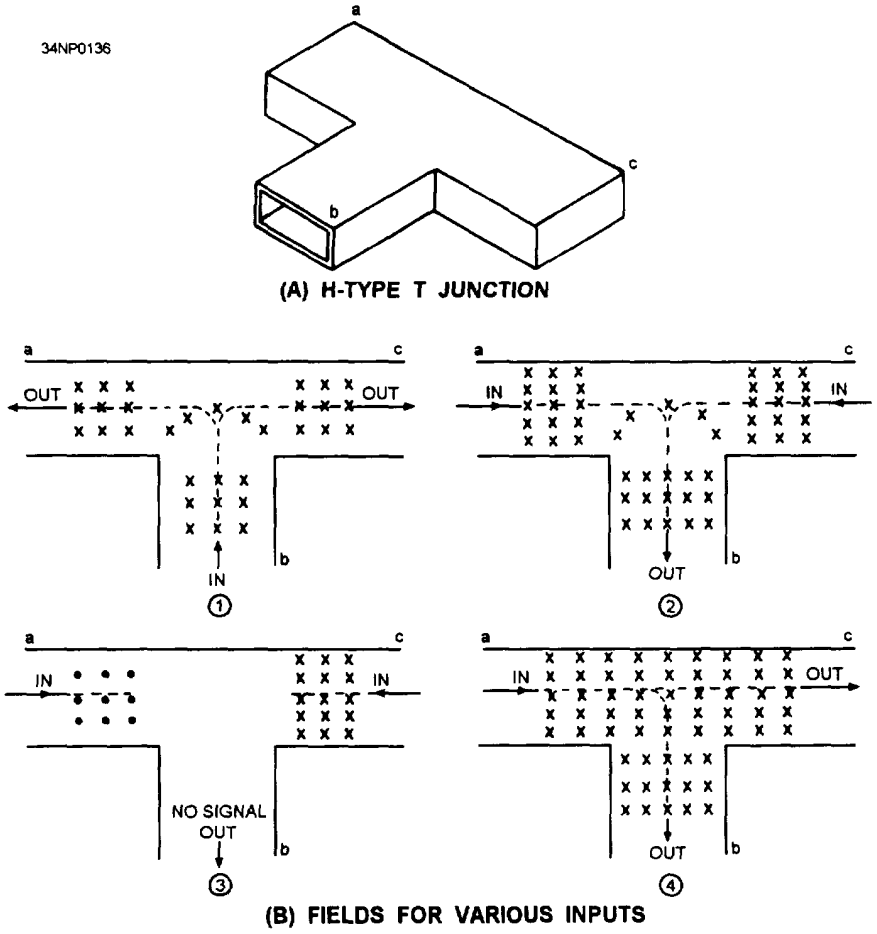


Figure 3-63.—E field in an H-type T junction.

In view 1 of figure 3-63, view B, the signal is fed into arm b and in-phase outputs are obtained from the a and c arms. In view 2, in-phase signals are fed into arms a and c and the output signal is obtained from the b arm because the fields add at the junction and induce E lines into the b arm. If 180-degree-out-of-phase signals are fed into arms a and c, as shown in view 3, no output is obtained from the b arm because the opposing fields cancel at the junction. If a signal is fed into the a arm, as shown in view 4 , outputs will be obtained from the b and c arms. The reverse is also true. If a signal is fed into the c arm, outputs will be obtained from the a and b arms.

MAGIC-T HYBRID JUNCTION.— A simplified version of the magic-T hybrid junction is shown in figure 3-64. The magic-T is a combination of the H-type and E-type T junctions. The most common application of this type of junction is as the mixer section for microwave radar receivers.

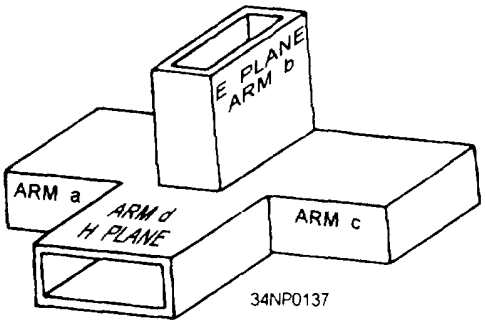


Figure 3-64.—Magic-T hybrid junction.

If a signal is fed into the b arm of the magic-T, it will divide into two out-of-phase components. As shown in figure 3-65, view A, these two components will move into the a and c arms. The signal entering the b arm will not enter the d arm because of the zero potential existing at the entrance of the d arm. The potential must be zero at this point to satisfy the boundary conditions of the b arm. This absence of potential is illustrated in views B and C where the magnitude of the E field in the b arm is indicated by the length of the arrows. Since the E lines are at maximum in the center of the b arm and minimum at the edge where the d arm entrance is located, no potential difference exists across the mouth of the d arm.

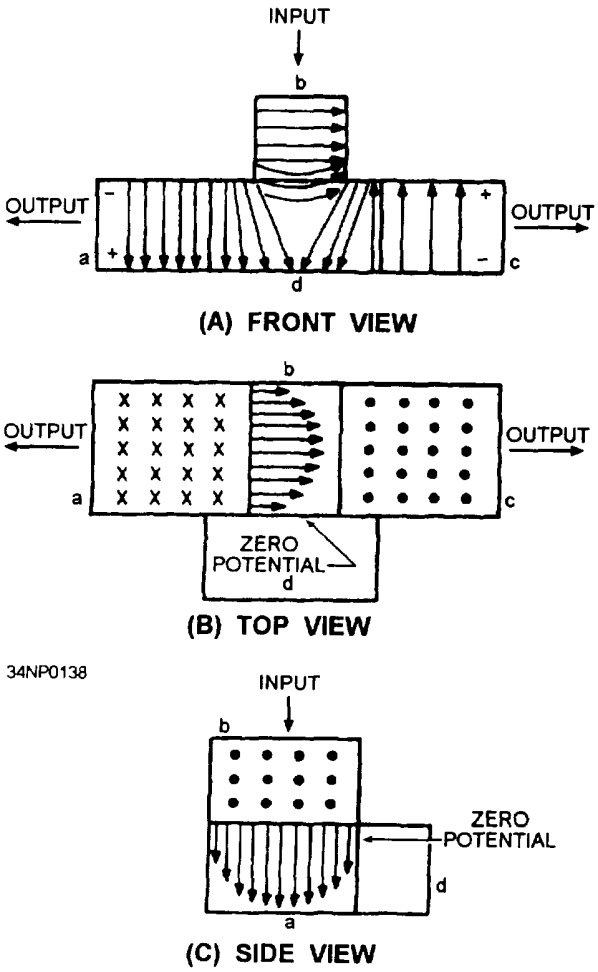


Figure 3-65.—Magic-T with input to arm b.

In summary, when an input is applied to arm b of the magic-T hybrid junction, the output signals from arms a and c are 180 degrees out of phase with each other, and no output occurs at the d arm.

The action that occurs when a signal is fed into the d arm of the magic-T is illustrated in figure 3-66. As with the H-type T junction, the signal entering the d arm divides and moves down the a and c arms as outputs that are in phase with each other and with the input. The shape of the E fields in motion is shown by the numbered curved slices. As the E field moves down the d arm, points 2 and 3 are at an equal potential. The energy divides equally into arms a and c, and the E fields in both arms become identical in shape. Since the potentials on both sides of the b arm are equal, no potential difference exists at the entrance to the b arm, resulting in no output.

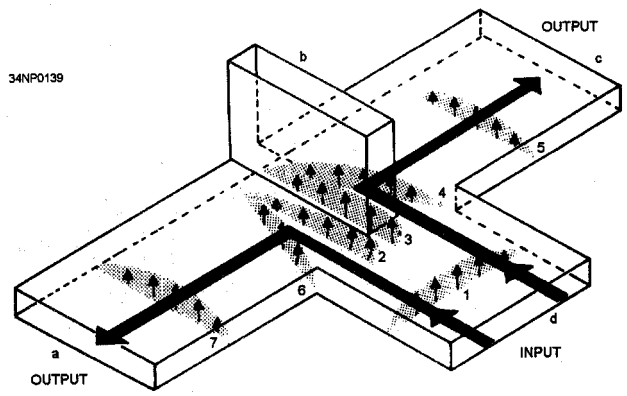


Figure 3-66.—Magic-T with input to arm d.

When an input signal is fed into the a arm as shown in figure 3-67, a portion of the energy is coupled into the b arm as it would be in an E-type T junction. An equal portion of the signal is coupled through the d arm because of the action of the H-type junction. The c arm has two fields across it that are out of phase with each other. Therefore, the fields cancel, resulting in no output at the c arm. The reverse of this action takes place if a signal is fed into the c arm, resulting in outputs at the b and d arms and no output at the a arm.

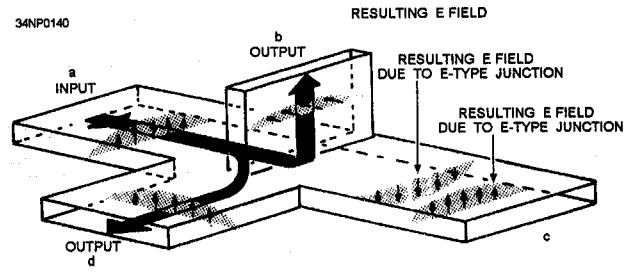


Figure 3-67.—Magic-T with input to arm a.

Unfortunately, when a signal is applied to any arm of a magic-T, the flow of energy in the output arms is affected by reflections. Reflections are caused by impedance mismatching at the junctions. These reflections are the cause of the two major disadvantages of the magic-T. First, the reflections represent a power loss since all the energy fed into the junction does not reach the load that the arms feed. Second, the reflections produce standing waves that can result in internal arcing. Thus, the maximum power a magic-T can handle is greatly reduced.

Reflections can be reduced by using some means of impedance matching that does not destroy the shape of the junctions. One method is shown in figure 3-68. A post is used to match the H plane, and an iris is used to match the E plane. Even though this method reduces reflections, it lowers the power-handling capability even further.

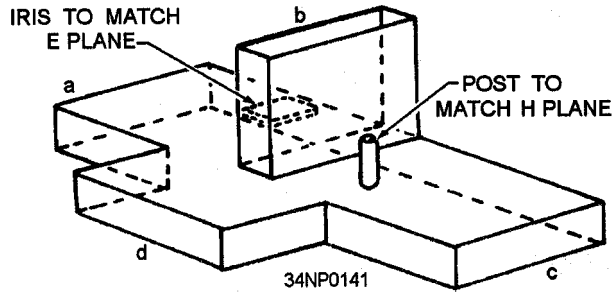


Figure 3-68.—Magic-T impedance matching.

HYBRID RING.— A type of hybrid junction that overcomes the power limitation of the magic-T is the hybrid ring, also called a RAT RACE. The hybrid ring, illustrated in figure 3-69, view A, is actually a modification of the magic-T. It is constructed of rectangular waveguides molded into a circular pattern. The arms are joined to the circular waveguide to form E-type T junctions. View B shows, in wavelengths, the dimensions required for a hybrid ring to operate properly.

The hybrid ring is used primarily in high-powered radar and communications systems to perform two functions. During the transmit period, the hybrid ring couples microwave energy from the transmitter to the antenna and allows no energy to reach the receiver. During the receive cycle, the hybrid ring couples energy from the antenna to the receiver and allows no energy to reach the transmitter. Any device that performs both of these functions is called a DUPLEXER. A duplexer permits a system to use the same antenna for both transmitting and receiving.

SUMMARY

This concludes our discussion on transmission lines and waveguides. In this volume you have been given a basic introduction on wave propagation from the time it leaves the transmitter to the point of reception. In volume 8 you will be introduced to a variety of electronic support systems.

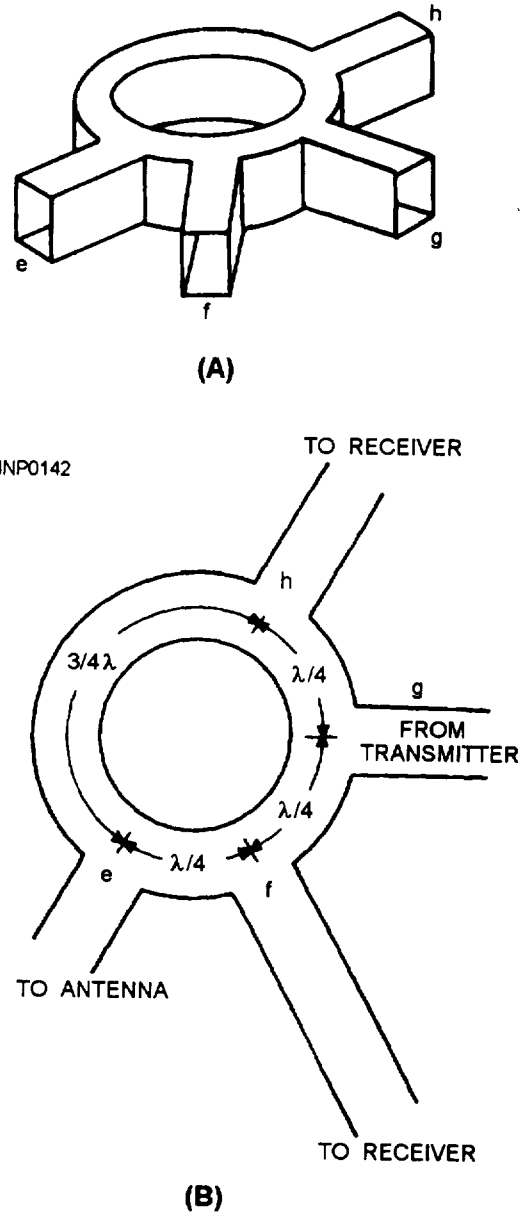


Figure 3-69.—Hybrid ring with wavelength measurements.

APPENDIX I

GLOSSARY

ABSORPTION—(1) Absorbing light waves. Does not allow any reflection or refraction; (2) Atmospheric absorption of rf energy with no reflection or refraction (adversely affects long-distance communications).

ACOUSTICS—The science of sound.

AMPLITUDE—The portion of a cycle measured from a reference line to a maximum value above (or to a maximum value below) the line.

ANGLE OF INCIDENCE—The angle between the incident wave and the normal.

ANGLE OF REFLECTION—The angle between the reflected wave and the normal.

ANGLE OF REFRACTION—The angle between the normal and the path of a wave through the second medium.

ANGSTROM UNIT—The unit used to define the wavelength of light waves.

ANISOTROPIC—The property of a radiator to emit strong radiation in one direction.

ANTENNA—A conductor or set of conductors used either to radiate rf energy into space or to collect rf energy from space.

APERTURE—See SLOT.

ARRAY OF ARRAYS—See COMBINATION ARRAY.

BAY—Part of an antenna array.

BEARING—An angular measurement that indicates the direction of an object in degrees from true north. Also called azimuth.

BEVERAGE ANTENNA—A horizontal, longwire antenna designed for reception and transmission of low-frequency, vertically polarized ground waves. Also known as WAVE ANTENNA.

BIDIRECTIONAL ARRAY—An array that radiates in opposite directions along the line of maximum radiation.

BROADSIDE ARRAY—An array in which the direction of maximum radiation is perpendicular to the plane containing the elements.

BOUNDARY CONDITIONS—The two conditions that the E-field and H-field within a waveguide must meet before energy will travel down the waveguide. The E-field must be perpendicular to the walls and the H-field must be in closed loops, parallel to the walls, and perpendicular to the E-field.

CAVITY RESONATOR—A space totally enclosed by a metallic conductor and supplied with energy in such a way that it becomes a source of electromagnetic oscillations. The size and shape of the enclosure determine the resonant frequency.

CENTER-FEED METHOD—Connecting the center of an antenna to a transmission line, which is then connected to the final (output) stage of the transmitter. Also known as CURRENT-FEED METHOD.

CHARACTERISTIC IMPEDANCE—The ratio of voltage to current at any given point on a transmission line. Represented by a value of impedance.

CHOKE JOINT—A joint between two sections of waveguide that provides a good electrical connection without power losses or reflections.

COAXIAL LINE—A type of transmission line that contains two concentric conductors.

COLLINEAR ARRAY—An array with all the elements in a straight line. Maximum radiation is perpendicular to the axis of the elements.

COMBINATION ARRAY—An array system that uses the characteristics of more than one array. Also known as ARRAY OF ARRAYS.

COMPLEX WAAE—A wave produced by combining two or more pure tones at the same time.

CONDUCTANCE—The opposite of resistance in transmission lines. The minute amount of resistance that is present in the insulator of a transmission line.

CONNECTED ARRAY—see DRIVEN ARRAY

COPPER LOSS—Power loss in copper conductors caused by the internal resistance of the conductors to current flow. Also know as I^2R LOSS.

CORNER-REFLECTOR ANTENNA—A half-wave antenna with a reflector consisting of two flat metal surfaces meeting at an angle behind the radiator.

COUNTERPOISE—A network of wire that is connected to a quarter-wave antenna at one end and provides the equivalent of an additional ¼ wavelength.

COUPLING DEVICE—A coupling coil that connects the transmitter to the feeder.

CREST (TOP)—The peak of the positive alternation (maximum value above the line) of a wave.

CRITICAL ANGLE—The maximum angle at which radio waves can be transmitted and still be refracted back to earth.

CRITICAL FREQUENCY—The maximum frequency at which a radio wave can be transmitted vertically and still be refracted back to earth.

CURRENT-FEED METHOD—See CENTER-FEED METHOD.

CURRENT STANDING-WAVE RATIO (ISWR)—The ratio of maximum to minimum current along a transmission line.

CUTOFF FREQUENCY—The frequency at which the attenuation of a waveguide increases sharply and below which a traveling wave in a given mode cannot be maintained. A frequency with a half wavelength that is greater than the wide dimension of a waveguide.

CYCLE—One complete alternation of a sine wave that has a maximum value above and a maximum value below the reference line.

DAMPING—Reduction of energy by absorption.

DENSITY—(1) The compactness of a substance; (2) Mass per unit volume.

DETECTOR—The device that responds to a wave or disturbance.

DIELECTRIC HEATING—The heating of an insulating material by placing it in a high frequency electric field.

DIELECTRIC LOSSES—The losses resulting from the heating effect on the dielectric material between conductors.

DIELECTRIC CONSTANT—The ratio of a given dielectric to the dielectric value of a vacuum.

DIFFRACTION—The bending of the paths of waves when the waves meet some form of obstruction.

DIPOLE—A common type of half-wave antenna made from a straight piece of wire cut in half. Each half operates at a quarter wavelength of the output.

DIRECTIONAL.—Radiation that varies with direction.

DIRECTIONAL COUPLER—A device that samples the energy traveling in a waveguide for use in another circuit.

DIRECTOR—The parasitic element of an array that reinforces energy coming from the driver toward itself.

DIRECTIVITY—The property of an array that causes more radiation to take place in certain directions than in others.

DISTRIBUTED CONSTANTS—The constants of inductance, capacitance, and resistance in a transmission line. The constants are spread along the entire length of the line and cannot be distinguished separately.

DOMINANT MODE—The easiest mode to produce in a waveguide, and also, the most efficient mode in terms of energy transfer.

DOPPLER EFFECT—The apparent change in frequency or pitch when a sound source moves either toward or away from a listener.

DOUBLET—Another name for the dipole antenna.

DRIVEN ARRAY—An array in which all of the elements are driven. Also known as CONNECTED ARRAY

DRIVEN ELEMENT—An element of an antenna (transmitting or receiving) that is connected directly to the transmission line.

DUMMY LOAD—A device used at the end of a transmission line or waveguide to convert transmitted energy into heat so no energy is radiated outward or reflected back.

E-FIELD—Electric field that exists when a difference in electrical potential causes a stress in the dielectric between two points. Also known as ELECTRIC FIELD.

E-TYPE T-JUNCTION—A waveguide junction in which the junction arm extends from the main waveguide in the same direction as the E-field in the waveguide.

ECHO—The reflection of the original sound wave as it bounces off a distant surface.

ELECTROMAGNETIC FIELD—The combination of an electric (E) field and a magnetic (H) field.

ELECTROMAGNETIC INTERFERENCE—Man-made or natural interference that degrades the quality of reception of radio waves.

ELECTROMAGNETIC RADIATION—The radiation of radio waves into space.

ELECTRIC FIELD—See E-FIELD.

ELEMENT—A part of an antenna that can be either an active radiator or a parasitic radiator.

END-FEED METHOD—Connecting one end of an antenna through a capacitor to the final output stage of a transmitter. Also known as VOLTAGE-FEED METHOD.

END-FIRE ARRAY—An array in which the direction of radiation is parallel to the axis of the array.

ELEVATION ANGLE—The angle between the line of sight to an object and the horizontal plane.

FADING—Variations in signal strength by atmospheric conditions.

FEEDER—A transmission line that carries energy to the antenna.

FLAT LINE—A transmission line that has no standing waves. This line requires no special tuning device to transfer maximum power.

FLEXIBLE COAXIAL LINE— coaxial line made with a flexible inner conductor insulated from the outer conductor by a solid, continuous insulating material.

FOLDED DIPOLE—An ordinary half-wave antenna (dipole) that has one or more additional conductors connected across the ends parallel to each other.

FOUR-ELEMENT ARRAY—An array with three parasitic elements and one driven element.

FREE-SPACE LOSS—The loss of energy of a radio wave because of the spreading of the wavefront as it travels from the transmitter.

FREQUENCY—The number of cycles that occur in one second. Usually expressed in Hertz.

FREQUENCY DIVERSITY—Transmitting (and receiving) of radio waves on two different frequencies simultaneously.

FRONT-TO-BACK RATIO—The ratio of the energy radiated in the principal direction to the energy radiated in the opposite direction.

FUNDAMENTAL FREQUENCY—The basic frequency or first harmonic frequency.

GAIN—The ratio between the amount of energy propagated from an antenna that is directional to the energy from the same antenna that would be propagated if the antenna were not directional.

GENERATOR END—See INPUT END

GROUND PLANE—The portion of a groundplane antenna that acts as ground.

GROUND-PLANE ANTENNA—A type of antenna that uses a ground plane as a simulated ground to produce low-angle radiation.

GROUND REFLECTION LOSS—The loss of rf energy each time a radio wave is reflected from the earth's surface.

GROUND SCREEN—A series of conductors buried below the surface of the earth and arranged in a radial pattern. Used to reduce losses in the ground.

GROUND WAVES—Radio waves that travel near the surface of the earth.

GROUP VELOCITY—The forward progress velocity of a wave front in a waveguide.

H-FIELD—Any space or region in which a magnetic force is exerted. The magnetic field may be produced by a current-carrying coil or conductor, by a permanent magnet, or by the earth itself. Also known as MAGNETIC FIELD.

H-TYPE T-JUNCTION—A waveguide junction in which the junction arm is parallel to the magnetic lines of force in the main waveguide.

HALF-WAVE DIPOLE ANTENNA—An antenna consisting of two rods (¼ wavelength h) in a straight line, that radiates electromagnetic energy.

HARMONIC—A frequency that is a whole number multiple of a smaller base frquency.

HERTZ ANTENNA—A half-wave antenna installed some distance above ground and positioned either vertically or horizontally.

HORN—A funnel-shaped section of waveguide used as a termination device and as a radiating antenna.

HORIZONTAL AXIS—On a graph, the straight line axis plotted from left to right.

HORIZONTAL PATTERN—The part of a radiation pattern that is radiated in all directions along the horizontal plane.

HORIZONTALLY POLARIZED—Waves that are radiated with their E-field component parallel to the earth's surface.

HYBRID JUNCTION—A waveguide junction that combines two or more basic T-junctions.

HYBRID RING—A hybrid-waveguide junction that combines a series of E-type T-junctions in a ring configuration.

I^2R LOSS—See COPPER LOSS.

INCIDENT WAVE—(1) The wave that strikes the surface of a medium; (2) The wave that travels from the sending end to the receiving end of a transmission line.

INDUCTION FIELD—The electromagnetic field produced about an antenna when current and voltage are present on the same antenna.

INDUCTION LOSSES—The losses that occur when the electromagnetic field around a conductor cuts through a nearby metallic object and induces a current into that object.

INPUT END—The end of a two-wire transmission line that is connected to a source. Also known as a GENERATOR END or a TRANSMITTER END.

INPUT IMPEDANCE—The impedance presented to the transmitter by the transmission line and its load.

INTERFERENCE—Any disturbance that produces an undesirable response or degrades a wave.

IONOSPHERE—The most important region of the atmosphere extending from 31 miles to 250 miles above the earth. Contains four cloud-like layers that affect radio waves.

IONOSPHERIC STORMS—Disturbances in the earth's magnetic field that make communications practical only at lower frequencies.

IONIZATION—The process of upsetting electrical neutrality.

IRIS—A metal plate with an opening through which electromagnetic waves may pass. Used as an impedance matching device in waveguides.

ISOTROPIC RADIATION—The radiation of energy equally in all directions.

LEAKAGE CURRENT—The small amount of current that flows between the conductors of a transmission line through the dielectric.

LOAD END—See OUTPUT END.

LOAD ISOLATOR—A passive attenuator in which the loss in one direction is much greater than that in the opposite direction. An example is a ferrite isolator for waveguides that allow energy to travel in only one direction.

LOADING—See LUMPED-IMPEDANCE TUNING.

LOBE—An area of a radiation pattern plotted on a polar-coordinate graph that represents maximum radiation.

LONG-WIRE ANTENNA—An antenna that is a wavelength or more long at its operating frequency.

LONGITUDINAL WAVES—Waves in which the disturbance (back and forth motion) takes place in the direction of propagation. Sometimes called compression waves.

LOOP—(1) The curves of a standing wave or antenna that represent amplitude of current or voltage; (2) A curved conductor that connects the ends of a coaxial cable or other transmission line and projects into a waveguide or resonant cavity for the purpose of injecting or extracting energy.

LOWEST USABLE FREQUENCY—The minimum operating frequency that can be used for communications between two points.

LUMPED CONSTANTS—The properties of inductance, capacitance, and resistance in a transmission line.

LUMPED-IMPEDANCE TUNING—The insertion of an inductor or capacitor in series with an antenna to lengthen or shorten the antenna electrically. Also known as LOADING.

LOOSE COUPLING—Inefficient coupling of energy from one circuit to another that is desirable in some applications. Also called weak coupling.

MAGIC-T JUNCTION—A combination of the H-type and E-type T-junctions.

MAGNETIC FIELD—See H-FIELD.

MAJOR LOBE—The lobe in which the greatest amount of radiation occurs.

MARCONI ANTENNA—A quarter-wave antenna oriented perpendicular to the earth and operated with one end grounded. Also known as QUARTER-WAVE ANTENNA.

MAXIMUM USABLE FREQUENCY—Maximum frequency that can be used for communications between two locations for a given time of day and a given angle of incidence.

MEDIUM—The substance through which a wave travels from one point to the next. Air, water, wood, etc., are examples of a medium.

METALLIC INSULATOR—A shorted quarter-wave section of transmission line.

MICROWAVE REGION—The portion of the electromagnetic spectrum from 1,000 megahertz to 100,000 megahertz.

MINOR LOBE—The lobe in which the radiation intensity is less than a major lobe.

MULTIELEMENT ARRAY—An array consisting of one or more arrays and classified as to directivity.

MULTIELEMENT PARASITIC ARRAY—An array that contains two or more parasitic elements and a driven element.

MULTIPATH—The multiple paths a radio wave may follow between transmitter and receiver.

NEGATIVE ALTERNATION—The portion of a sine wave below the reference line.

NODE—The fixed minimum points of voltage or current on a standing wave or antenna.

NONDIRECTIONAL—See OMNIDIRECTIONAL,

NONRESONANT LINE—A transmission line that has no standing waves of current or voltage.

NORMAL—The imaginary line perpendicular to the point at which the incident wave strikes the reflecting surface. Also called the perpendicular.

NULL—On a polar-coordinate graph, the area that represents minimum or 0 radiation.

OMNIDIRECTIONAL—Transmitting in all directions. Also known as NONDIRECTIONAL.

OPEN-ENDED LINE—A transmission line that has an infinitely large terminating impedance.

OPTIMUM WORKING FREQUENCY—The most practical operating frequency that can be used with the least amount of problems; roughly 85 percent of the maximum usable frequency.

ORIGIN—The point on a graph where the vertical and horizontal axes cross each other.

OUTPUT END—The end of a transmission line that is opposite the source. Also known as RECEIVING END.

OUTPUT IMPEDANCE—The impedance presented to the load by the transmission line and its source.

PARALLEL RESONANT CIRCUIT—A circuit that acts as a high impedance at resonance.

PARALLEL-WIRE—A type of transmission line consisting of two parallel wires.

PARASITIC ARRAY—An array that has one or more parasitic elements.

PARASITIC ELEMENT—The passive element of an antenna array that is connected to neither the transmission line nor the driven element.

PERIOD—The amount of time required for completion of one full cycle.

PHASE SHIFTER—A device used to change the phase relationship between two ac signals.

PLANE OF POLARIZATION—The plane (vertical or horizontal) with respect to the earth in which the E-field propagates.

POSITIVE ALTERNATION—The portion of a sine wave above the reference line.

POWER GAIN—The ratio of the radiated power of an antenna compared to the output power of a standard antenna. A measure of antenna efficiency usually expressed in decibels. Also referred to as POWER RATIO.

POWER LOSS—The heat loss in a conductor as current flows through it.

POWER RATIO—See POWER GAIN.

POWER STANDING—WAVE RATIO (PSWR)—The ratio of the square of the maximum and minimum voltages of a transmission line.

PROPAGATION—Waves traveling through a medium.

PROBE—A metal rod that projects into, but is insulated from, a waveguide or resonant cavity and used to inject or extract energy.

QUARTER-WAVE ANTENNA—See MARCONI ANTENNA.

RADIATION FIELD—The electromagnetic field that detaches itself from an antenna and travels through space.

RADIATION LOSSES—The losses that occur when magnetic lines of force about a conductor are projected into space as radiation and are not returned to the conductor as the cycle alternates.

RADIATION PATTERN—A plot of the radiated energy from an antenna.

RADIATION RESISTANCE—The resistance, which if inserted in place of an antenna, would consume the same amount of power as that radiated by the antenna.

RADIO FREQUENCIES—Electromagnetic frequencies that fall between 3 kilohertz and 300 gigahertz and are used for radio communications.

RADIO HORIZON—The boundary beyond the natural horizon in which radio waves cannot be propagated over the earth's surface.

RADIO WAVE—(1) A form of radiant energy that can neither be seen nor felt; (2) An electromagnetic wave generated by a transmitter.

RAREFIED WAVE—A longitudinal wave that has been expanded or rarefied (made less dense) as it moves away from the source.

RECEIVER—The object that responds to a wave or disturbance. Same as detector.

RECEIVING ANTENNA—The device used to pick up an rf signal from space.

RECEIVING END—See OUTPUT END.

RECIPROCITY—The ability of an antenna to both transmit and receive electromagnetic energy with equal efficiency.

REFLECTED WAVE—(1) The wave that reflects back from a medium; (2) Waves traveling from the load back to the generator on a transmission line; (3) The wave moving back to the sending end of a transmission line after reflection has occurred.

REFLECTION WAVES—Waves that are neither transmitted nor absorbed, but are reflected from the surface of the medium they encounter.

REFLECTOR—The parasitic element of an array that causes maximum energy radiation in a direction toward the driven element.

REFRACTION—The changing of direction as a wave leaves one medium and enters another medium of a different density.

REFRACTIVE INDEX—The ratio of the phase velocity of a wave in free space to the phase velocity of the wave in a given substance (dielectric).

RERADIATION—The reception and retransmission of radio waves caused by turbulence in the troposphere.

RESONANCE—The condition produced when the frequency of vibrations are the same as the natural frequency (of a cavity), The vibrations reinforce each other.

RESONANT LINE—A transmission line that has standing waves of current and voltage.

RHOMBIC ANTENNA—A diamond-shaped antenna used widely for long-distance, high-frequency transmission and reception.

RIGID COAXIAL LINE—A coxial line consisting of a central, insulated wire (inner conductor) mounted inside a tubular outer conductor.

ROTATING JOINT—A joint that permits one section of a transmission line or waveguide to rotate continuously with respect to another while passing energy through the joint. Also called a rotary coupler.

SCATTER ANGLE—The angle at which the receiving antenna must be aimed to capture the scattered energy of tropospheric scatter.

SELF-INDUCTION—The phenomenon caused by the expanding and collapsing fields of an electron that encircles other electrons and retards the movement of the encircled electrons.

SERIES RESONANT CIRCUIT—A circuit that acts as a low impedance at resonance.

SHIELDED PAIR—A line consisting of parallel conductors separated from each other and surrounded by a solid dielectric.

SHORT-CIRCUITED LINE—A transmission line that has a terminating impedance equal to 0.

SKIN EFFECT—The tendency for alternating current to concentrate in the surface layer of a conductor. The effect increases with frequency and serves to increase the effective resistance of the conductor.

SKIP DISTANCE—The distance from a transmitter to the point where the sky wave is first returned to earth.

SKIP ZONE—A zone of silence between the point where the ground wave becomes too weak for reception and the point where the sky wave is first returned to earth.

SKY WAVES—Radio waves reflected back to earth from the ionosphere.

SLOT—Narrow opening in a waveguide wall used to couple energy in or out of the waveguide. Also called an APERTURE or a WINDOW.

SOURCE—(1) The object that produces waves or disturbance; (2) The name given to the end of a two-wire transmission line that is connected to a source.

SPACE DIVERSITY—Reception of radio waves by two or more antennas spaced some distance apart,

SPACE WAVE—A radio wave that travels directly from the transmitter to the receiver and remains in the troposphere.

SPECTRUM—(1) The entire range of electromagnetic waves; (2) VISIBLE. The range of electromagnetic waves that stimulate the sense of sight;

(3) ELECTROMAGNETIC. The entire range of electromagnetic waves arranged in order of their frequencies.

SPORADIC E LAYER—Irregular cloud-like patches of unusually high ionization. Often forms at heights near the normal E-layer.

SPREADER—Insulator used with transmission lines and antennas to keep the parallel wires separated.

STANDING WAVE—The distribution of voltage and current formed by the incident and reflected waves, which have minimum and maximum points on a resultant wave that appears to stand still.

STANDING-WAVE RATIO (SWR)—The ratio of the maximum to the minimum amplitudes of corresponding components of a field, voltage, or current along a transmission line or waveguide in the direction of propagation measured at a given frequency. Measures the perfection of the termination of the line.

STRATOSPHERE—Located between the troposphere and the ionosphere. Has little effect on radio waves.

STUB—Short section of a transmission line used to match the impedance of a transmission line to an antenna. Can also be used to produce desired phase relationships between connected elements of an antenna.

SUDDEN IONOSPHERIC DISTURBANCE—An irregular ionospheric disturbance that can totally blank out hf radio communications.

SURFACE WAVE—A radio wave that travels along the contours of the earth, thereby being highly attenuated.

TEMPERATURE INVERSION—The condition in which warm air is formed above a layer of cool air that is near the earth's surface.

THREE-ELEMENT ARRAY—An array with two parasitic elements (reflector and director) and a driven element.

TRANSMISSION LINE—A device designed to guide electrical energy from one point to another.

TRANSMITTING ANTENNA—The device used to send the transmitted signal energy into space.

TRANSMISSION MEDIUMS—The various types of lines and waveguides used as transmission lines.

TRANSMITTER END—See INPUT END.

TRANSVERSE WAVE MOTION—The up and down motion of a wave as the wave moves outward.

TRANSVERSE ELECTRIC MODE—The entire electric field in a waveguide is perpendicular to the wide dimension and the magnetic field is parallel to the length. Also called the TE mode.

TRANSVERSE MAGNETIC MODE—The entire magnetic field in a waveguide is perpendicular to the wide dimension ("a" wall) and some portion of the electric field is parallel to the length. Also called the TM mode.

TROPOSPHERE—The portion of the atmosphere closest to the earth's surface, where all weather phenomena take place.

TROPOSPHERIC SCATTER—The propagation of radio waves in the troposphere by means of scatter.

TROUGH (BOTTOM)—The peak of the negative alternation (maximum value below the line).

TUNED LINE—Another name for the resonant line. This line uses tuning devices to eliminate the reactance and to transfer maximum power from the source to the line.

TURNSTILE ANTENNA—A type of antenna used in vhf communications that is omnidirectional

and consists of two horizontal half-wave antennas mounted at right angles to each other in the same horizontal plane.

TWISTED PAIR—A line consisting of two insulated wires twisted together to form a flexible line without the use of spacers.

TWO-WIRE OPEN LINE—A parallel line consisting of two wires that are generally spaced from 2 to 6 inches apart by insulating spacers.

TWO-WIRE RIBBON (TWIN LEAD)—A parallel line similar to a two-wire open line except that uniform spacing is assured by embedding the two wires in a low-loss dielectric.

UNIDIRECTIONAL ARRAY—An array that radiates in only one general direction.

UNTUNED LINE—Another name for the flat or nonresonant line.

V ANTENNA—A bidirectional antenna, shaped like a V, which is widely used for communications.

VELOCITY—The rate at which a disturbance travels through a medium.

VERTICAL AXIS—On a graph, the straight line axis oriented from bottom to top.

VERTICAL PATTERN—The part of a radiation pattern that is radiated in the vertical plane.

VERTICAL PLANE—An imaginary plane that is perpendicular to the horizontal plane.

VERTICALLY POLARIZED—Waves radiated with the E-field component perpendicular to the earth's surface.

VOLTAGE-FEED METHOD—See END-FEED METHOD.

VOLTAGE STANDING-WAVE RATIO (VSWR)—The ratio of maximum to minimum voltage of a transmission line.

WAVE ANTENNA—See BEVERAGE ANTENNA.

WAVE MOTION—A recurring disturbance advancing through space with or without the use of a physical medium.

WAVE TRAIN—A continuous series of waves with the same amplitude and wavelength.

WAVEFRONT—A small section of an expanding sphere of electromagnetic radiation, perpendicular to the direction of travel of the energy.

WAVEGUIDE—A rectangular, circular, or elliptical metal pipe designed to transport electro-magnetic waves through its interior.

WAVEGUIDE MODE OF OPERATION—Particular field configuration in a waveguide that satisfies the boundary conditions. Usually divided into two broad types: the transverse electric (TE) and the transverse magnetic (TM).

WAVEGUIDE POSTS—A rod of conductive material used as impedance-changing devices in waveguides.

WAVEGUIDE SCREW—A screw that projects into a waveguide for the purpose of changing the impedance.

WAVELENGTH—(1) The distance in space occupied by 1 cycle of a radio wave at any given instant; (2) The distance a disturbance travels during one period of vibration.

WINDOW—See Slot.

YAGI ANTENNA—A multielement parasitic array. Elements lie in the same plane as those of the end-fire array.

APPENDIX II

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Assignment Questions

Information: The text pages that you are to study are provided at the beginning of the assignment questions.

ASSIGNMENT 1

Textbook Assignment: "Wave Propagation," chapter 1, pages 1-1 through 1-14.

1-1. Which of the following factors can affect atmospheric conditions?

1. Geographic height
2. Geographic location
3. Changes in time
4. All of the above

1-2. In what portion of the atmosphere does the majority of weather phenomena take place?

1. Ionosphere
2. Stratosphere
3. Troposphere
4. Hydrosphere

1-3. Because the stratosphere is a relatively calm region with little or no temperature change, it will have almost no effect on radio wave propagation.

1. True
2. False

1-4. Variations in the ionosphere resulting from changes in the sun's activity are known as

1. regular variations
2. irregular variations
3. both 1 and 2 above
4. seasons

1-5. The regular variations in the ionosphere can be separated into how many classes?

1. One
2. Two
3. Three
4. Four

1-6. In ionization, when an electron is knocked free from a neutral gas atom, what is the overall charge of the atom?

1. Negative
2. Positive
3. Neutral
4. Inverted

1-7. The frequency of ultraviolet light passing through the atmosphere has what relationship to the ionospheric layer it ionizes?

1. It is inversely proportional
2. It is directly proportional
3. It is inversely proportional during the day and directly proportional at night
4. It is directly proportional during the day and inversely proportional at night

1-8. What term best describes the process that returns positive ions to their original neutral state?

1. Refraction
2. Recombination
3. Ionization
4. Polarization

1-9. At what approximate time of day is the density of the ionospheric layers at its lowest level?

1. Just before sunrise
2. Mid-morning
3. Afternoon
4. Sunset

1-10. How many distinct layers make up the ionosphere?

1. One
2. Two
3. Three
4. Four

1-11. At what frequencies does the combination of the earth's surface and the D layer act as a waveguide?

1. Vlf
2. Lf
3. Mf
4. Hf

1-12. The D layer loses its absorptive qualities at frequencies above what level?

1. 30 MHz
2. 20 MHz
3. 10 MHz
4. 3 MHz

1-13. What is the approximate range of the E layer above the earth's surface?

1. 30–54 miles
2. 55–90 miles
3. 91–130 miles
4. 131–160 miles

1-14. Frequencies above what level pass through the E layer unaffected?

1. 50 MHz
2. 100 MHz
3. 150 MHz
4. 200 MHz

1-15. During daylight hours, the F layer will divide into how many separate layers?

1. Five
2. Two
3. Three
4. Four

1-16. Most high-frequency, long-range communications occur in what layer(s) of the ionosphere?

1. D
2. E
3. F
4. H

1-17. Which of the following is NOT a factor for radio wave refraction?

1. Ionization density of the layer
2. Frequency of the radio wave
3. Angle of incidence
4. Transmitter power

1-18. For any given ionized layer, the critical frequency is just below the escape point.

1. True
2. False

1-19. The critical angle for radio wave propagation depends on what two factors?

1. Angle of incidence and layer density only
2. Layer density and wavelength only
3. Angle of incidence and wavelength only
4. Wavelength and antenna height only

1-20. What term best describes the area located between the transmitting antenna and the point where the sky wave first returns to the earth?

1. Ground wave
2. Skip zone
3. Skip distance
4. Ace area

1-21. Which of the following factors will affect the outer limits of the skip zone?

1. Frequency
2. Sunspot activity
3. Angle of transmission
4. All of the above

1-22. Radio waves reflecting from the earth's surface or the ionosphere, 180 degrees out of phase, have what effect, if any, at the receiving station?

1. The signal will be weak or faded
2. The signal will be stronger
3. The signal will be garbled
4. None

1-23. For ionospheric reflection to occur, the ionized layer must not be thicker than how many wavelengths of the transmitted frequency?

1. One
2. Two
3. Three
4. Four

1-24. The ability of radio waves to turn sharp corners and bend around obstacles is known as

1. reflection
2. refraction
3. diffraction
4. waveshaping

1-25. Which of the following definitions best describes a shadow zone?

1. The area of complete coverage at vlf frequencies
2. The area within the diameter of an obstruction
3. The area ranging the height of the obstruction
4. The area on the opposite side of the obstruction, in line-of-site from the transmitter to the receiver

1-26. What type of fading occurs for the longest amount of time?

1. Phase shift
2. Absorption
3. Multipath
4. Diffraction

1-27. Which of the following are examples of multipath radio wave transmissions?

1. Groundwaves
2. Ionospheric refractions
3. Reflection from the earth's surface
4. All of the above

1-28. Fading on the majority of the ionospheric circuits is a result of what particular type of fading?

1. Selective
2. Absorption
3. Multipath
4. Weather

IN ANSWERING QUESTIONS 1-29 AND 1-30, SELECT FROM THE FOLLOWING LIST THE DEFINITION OF THE INDICATED TERM.

A. Two or more receiving antennas spaced apart to produce a usable signal

B. Two or more receiving antennas of varying heights located together

C. The use of two separate transmitters and receivers on different frequencies transmitting the same information

D. The use of two separate transmitters and receivers on the same frequency transmitting the same information

1-29. Space diversity.

1. A
2. B
3. C
4. D

1-30. Frequency diversity.

1. A
2. B
3. C
4. D

1-31. A wide band of frequencies is transmitted and selective fading occurs. Which of the following statements best describes the effect of the fading on the signal?

1. It affects various frequencies
2. It can cause changes in phase and amplitude
3. It can cause severe distortion and limit total signal strength
4. All of the above

1-32. Which ionospheric layer is most dense during the winter?

1. E
2. D
3. F2
4. F1

1-33. During the 27-day sunspot cycle, which ionospheric layer experiences the greatest fluctuations in density?

1. D
2. E
3. F1
4. F2

IN ANSWERING QUESTIONS 1-34 THROUGH 1-38, SELECT FROM THE FOLLOWING LIST THE DEFINITION OF THE INDICATED TERM.

A. Depends on the angle of the sun; refracts hf waves during the day, up to 20 MHz, to distances of 1200 miles; greatly reduced at night

B. Reflects vlf waves for long-range communications; refracts lf and mf for short-range communications; has little effect on vhf and above; gone at night

C. Density depends on the angle of the sun; its main effect is absorption of hf waves passing through to the F2 layer

D. Provides long-range hf communications; very variable; height and density change with time of day, season, and sunspot activity

E. Structure and density depend on the time of day and the angle of the sun; consists of one layer at night and two layers during the day

1-34. D layer.

1. A
2. B
3. C
4. D

1-35. E layer.

1. A
2. B
3. C
4. E

1-36. F layer.

1. B
2. C
3. D
4. E

1-37. F1 layer.

1. A
2. B
3. C
4. D

1-38. F2 layer.

1. A
2. C
3. D
4. E

1-39. During periods of maximum sunspot activity within the eleven-year cycle, critical frequencies for all layers increase.

1. True
2. False

1-40. Which of the following problems is NOT a negative side effect of the sporadic E layer?

1. Causes increased multipath problems
2. Provides additional absorption
3. Blanks out more favorable layers
4. Increased static in line of sight communications

1-41. When sudden ionospheric disturbances (SID) occurs, which ionospheric layer is affected the most?

1. D
2. E
3. F1
4. F2

1-42. What effect do ionospheric storms have on (a) the range of frequencies and (b) the working frequency used for communications?

1. (a) Increase (b) increase
2. (a) Decrease (b) decrease
3. (a) Increase (b) decrease
4. (a) Decrease (b) increase

1-43. What form of precipitation has the greatest absorption effect on RF energy?

1. Fog
2. Snow
3. Rain
4. Hail

1-44. The duct effect produced by temperature inversion allows for long-distance communications over what frequency band?

1. Vlf
2. Lf
3. Hf
4. Vhf

1-45. Which of the following factors affect(s) the amount of ground reflection loss when a radio wave is reflected from the earth's surface?

1. Angle of incidence
2. Ground irregularities
3. Electrical conductivity at the point of reflection
4. All of the above

1-46. As an RF wave increases in distance, the wavefront spreads out, reducing the amount of energy available within any given unit of area. This action produces what type of energy loss?

1. Absorption
2. Ground reflection
3. Freespace
4. Spread

1-47. Radio waves above the MUF will experience what effect when refracted from the ionosphere?

1. They will fall short of the desired location
2. They will overshoot the desired location
3. They will be absorbed by lower layers
4. They will experience multipath fading

1-48. Variations in the ionosphere may change a preexisting muf. This is especially true because of the volatility of which of the following layers?

1. F1
2. F2
3. D
4. E

1-49. Radio waves that are propagated below the LUF are affected by what problem(s)?

1. Increased absorption
2. Higher levels of atmospheric noise
3. Higher rate of refraction
4. All of the above

1-50. The frequency that will avoid the problems of multipath fading, absorption, noise, and rapid changes in the ionosphere is known by what term?

1. LUF
2. MUF
3. FOT
4. LOS

ASSIGNMENT 2

Textbook Assignment: "Antennas," chapter 2, pages 2-1 through 2-32.

2-1. Electromagnetic radiation from an antenna is made up of what two components?

1. E and H fields
2. Ground and sky waves
3. Vertical and horizontal wavefronts
4. Reflected and refracted energy

2-2. What determines the size of a transmitting antenna?

1. Transmitter power
2. Available space
3. Operating frequency
4. Distance to be transmitted

2-3. Most practical transmitting antennas are divided into two classifications, Hertz and Marconi.

1. T
2. F

2-4. Hertz antennas are designed to operate at what wavelength in relationship to their operating frequency?

1. Quarter-wave
2. Half-wave
3. Three quarter-wave
4. Full-wave

2-5. Marconi antennas are used for operating frequencies below what level?

1. 10 MHz
2. 6 MHz
3. 4 MHz
4. 2 MHz

2-6. All antennas regardless of their shape or size have how many basic characteristics?

1. 1
2. 2
3. 3
4. 4

2-7. The ability to use the same antenna for both transmitting and receiving is known by what term?

1. Gain
2. Reciprocity
3. Directivity
4. Polarization

2-8. The ability of an antenna or array to focus energy in one or more specific directions is represented by a measurement of what antenna property?

1. Signal Strength
2. Reciprocity
3. Directivity
4. Polarization

2-9. The gain of a transmitting antenna is 9 dB, what will the gain be for the same antenna used for receiving?

1. 9 dB
2. 6 dB
3. 4 dB
4. 3 dB

2-10. Which, if any, of the following components of a radiated electromagnetic field determines its direction of polarization?

1. H lines
2. E lines
3. Angle of Propagation
4. None of the above

2-11. Over long distances the polarization of a radiated wave changes, at what frequencies will this change be the most dramatic?

1. VLF
2. LF
3. MF
4. HF

2-12. A transmitting antenna at ground level should be polarized in what manner to achieve best signal strength?

1. Horizontally
2. Vertically
3. Circularly
4. Linearly

2-13. What term describes the distance a wave travels during the period of one cycle?

1. Wavelength
2. Frequency
3. Travel time
4. Radiation rate

IN ANSWERING QUESTIONS 2-14 AND 2-15, REFER TO FIGURE 2-4 OF THE TEXT.

2-14. The points of high current and voltage are best described by which of the following terms?

1. Peaks
2. Crescents
3. Loops
4. Highs

2-15. The points of minimum voltage and minimum current are represented by which of the following terms?

1. Lows
2. Valleys
3. Descents
4. Nodes

2-16. An antenna at resonance will transmit at maximum efficiency; an antenna that is not at resonance will lose power in which of the following ways?

1. Skin effect loss
2. Heat loss
3. Ground absorption
4. Wave scattering

2-17. An antenna that radiates energy in all directions is said to have what type of radiation pattern?

1. Isotropic
2. Anisotropic
3. Bysotropic
4. Circumstropic

2-18. An antenna that radiates energy more strongly in one direction than another is said to have what type of radiation pattern?

1. Isotropic
2. Anisotropic
3. Bysotropic
4. Circumstropic

2-19. When viewing a radiation pattern graph, you can expect the areas of maximum and minimum radiation be identified by which of the following terms?

1. High and low probes
2. Maximum and minimum points
3. Major and minor lobes
4. Positive and negative lobes

2-20. If an antenna is too short for the wavelength being used, what electrical compensation must be introduced for the antenna to achieve resonance?

1. Lumped resistance
2. Lumped capacitive reactance
3. Lumped inductive reactance
4. More power

2-21. If an antenna is too long for the wavelength being used, what electrical compensation must be introduced for the antenna to achieve resonance?

1. Lumped resistance
2. Lumped capacitive reactance
3. Lumped inductive reactance
4. Less power

2-22. A ground screen is a series of conductors buried 1 or 2 feet below the surface in a radial pattern and is usually of what length in comparison to the wavelength being used?

1. One-quarter wavelength
2. One-half wavelength
3. Three-quarter wavelength
4. Full wavelength

2-23. When would a counterpoise be used?

1. When easy access to the antenna base is necessary
2. When the surface below is solid rock
3. When the surface below is sandy ground
4. All the above

2-24. Capacitive top-loading helps to increase which of the following antenna characteristics?

1. Bandwidth
2. Power-handling
3. Directivity
4. Radiation efficiency

2-25. What is the most limiting characteristic of the Yagi antenna?

1. Power-handling
2. Narrow bandwidth
3. Physical size
4. Lack of directivity

2-26. In general, log-periodic antennas have which of the following characteristics?

1. Medium power handling capabilities
2. High gain
3. Extremely broad bandwidth
4. All the above

2-27. A typical vertical monopole log periodic antenna designed to cover a frequency range of 2 to 30 MHz will require approximately how many acres of land for its ground plane system?

1. 1 acre
2. 2 acres
3. 3 acres
4. 4 acres

2-28. A sector log-periodic array can act as an antenna for a minimum of what number of transmit or receive systems?

1. 1
2. 2
3. 3
4. 4

2-29. The most distinct advantage of the rotatable log-periodic antenna is its ability to perform what function?

1. Rotate 360 degrees
2. Rotate from horizontal to vertical and back
3. Ability to handle high transmitter power
4. Ability to produce high antenna gain

2-30. What is the average power handling capability of an Inverted Cone antenna?

1. 20 kw
2. 30 kw
3. 40 kw
4. 50 kw

2-31. What determines the gain and directivity of a Rhombic antenna?

1. Transmitter power
2. Antenna height
3. Radiated wave interaction
4. Transmitted frequency

2-32. Most Whip antennas require some kind of a tuning system to improve bandwidth and power handling capabilities.

1. T
2. F

2-33. Why are UHF and VHF antennas on board ship installed as high as possible?

1. To prevent radiation hazard to personnel
2. To prevent radiation hazard to ordinance
3. To increase power handling capabilities
4. To prevent unwanted directivity in the radiation pattern from mast structures

2-34. The central feed section for both the biconical and center-fed dipole are protected by what type of covering?

1. SCOTCHCOAT
2. RTV
3. Laminated fiberglass
4. Rubber shield

2-35. The adjustable stub on the AS-390/SRC uhf antenna is used to adjust what antenna characteristic?

1. The counterpoise angle
2. The input impedance
3. The radiation angle
4. The feed point

2-36. The OE-82B/WSC-I(V) antenna group uses what type of polarization?

1. Vertical
2. Horizontal
3. Right-hand circular
4. Left-hand circular

2-37. The AN/WSC-5 (V) shore station antenna consists of what number of OE-82A/WSC-1 (V) assemblies?

1. 1
2. 2
3. 3
4. 4

2-38. What does the acronym LPI stand for?

1. Low power interference
2. Low probability of intercept
3. Low phase intercept
4. Last pass intercept

2-39. The reflectors for the AN/WSC-6 (V) are mounted on three-axis pedestals and provide auto tracking using what scanning technique?

1. Conical
2. Peripheral
3. Vertical
4. Horizontal

2-40. Antenna tuning is accomplished using what piece or pieces of equipment?

1. Couplers
2. Tuners
3. Multicouplers
4. All the above

2-41. Antenna multicouplers are used to match more than one transmitter or receiver to what number of antennas?

1. 1
2. 2
3. 3
4. 4

2-42. The AN/URA-38 antenna coupler is an automatic tuning system primarily used with which radio transmitter?

1. AN/WSC-3
2. AN/URT-23
3. AN/URC-80
4. AN/FRT-84

2-43. The AN/SRA-57 coupler group operates in which of the following frequency ranges?

1. 2- 6 Mhz
2. 4-12 Mhz
3. 10-30 Mhz
4. 40-60 Mhz

2-44. How many channels are provided with the AN/SRA-12 multicoupler?

1. 4
2. 5
3. 6
4. 7

2-45. What type of radar would use a truncated paraboloid reflector that has been rotated 90 degrees?

1. A surface search
2. An air search
3. A navigation
4. A height-finder

2-46. Of the following methods, which is NOT used to feed a cylindrical paraboloid reflector?

1. A linear array of dipoles
2. A slit in the side of a waveguide
3. A thin waveguide radiator
4. A quarter-wave stub

2-47. The elements of a broadside array are spaced one-half wavelength apart and are spaced how many wavelengths away from the reflector?

1. One-eighth
2. One-quarter
3. One-half
4. Three-quarter

2-48. What is the advantage, if any, to offsetting a feedhorn radiator for a parabolic dish?

1. A broader beam angle
2. The elimination of shadows
3. A narrower beam angle
4. No advantage

2-49. What is the range in nautical miles of the AN/GPN-27 radar?

1. 55
2. 75
3. 105
4. 155

2-50. What is the purpose of the jackscrew on the AS-3263/SPS-49(V) antenna?

1. To adjust the beam width
2. To vary the antenna feed horn focal distance
3. To adjust the beam elevation angle
4. To lockdown the antenna for PM

2-51. The OE-172/SPS-55 antenna normally operates in the linearly polarized mode, for what reason would you use the circular polarized mode?

1. To compensate for the ships pitch and roll
2. To prevent jamming
3. To reduce return echoes from precipitation
4. To achieve over the horizon coverage

2-52. Which of the following is NOT a mode of operation for the AN/SPN-35A radar set?

1. Final
2. Dual
3. Surveillance
4. Simultaneous

2-53. The two primary safety concerns associated with rf fields are rf burns and injuries caused by dielectric heating.

1. T
2. F

2-54. When a person is standing in an rf field, power in excess of what level will cause a noticeable rise in body temperature?

1. 5 milliwatts
2. 10 milliwatts
3. 15 milliwatts
4. 20 milliwatts

2-55. When working aloft, what safety precaution(s) must be followed?

1. Tag out the antenna at the switchboard to prevent it from becoming operational
2. Secure motor safety switches for rotating antennas
3. Wear the proper oxygen breathing apparatus when working near a stack
4. All the above

ASSIGNMENT 3

Textbook Assignment: "Transmission Lines and Waveguides," chapter 3, pages 3-1 through 3-32.

3-1. A transmission line is designed to perform which of the following functions?

1. Disperse energy in all directions
2. Detune a transmitter to match the load
3. Guide electrical energy from point to point
4. Replace the antenna in a communications system

3-2. The conductance value of a transmission line represents which of the following values?

1. Expected value of current flow through the insulation
2. Expected value of voltage supplied by the transmitter
3. Value of the lump and distributed constants of the line divided by impedance
4. Value of the lumped constants of the line as seen by the source and the load

3-3. Distributed constants in a transmission line are distributed in which of the following ways?

1. Between ground and any single point on the line
2. Along the length of the line
3. According to the thickness of the line
4. According to the cross-sectional area of the line

3-4. Leakage current in a two-wire transmission line is the current that flows through what component?

1. The resistor
2. The inductor
3. The insulator
4. The conductor

3-5. Conductance is the reciprocal of what electrical property?

1. Inductance
2. Resistance
3. Capacitance
4. Reciprocity

3-6. A transmission line that has current flowing through it has which of the following fields about it?

1. Electric only
2. Magnetic only
3. Both electric and magnetic
4. Capacitive

3-7. A measurement of the voltage to current ratio (E_{in}/I_{out}) at the input end of a transmission line is called the

1. input-gain rate
2. voltage-gain ratio
3. output impedance
4. input impedance

3-8. The characteristic impedance (2.) of a transmission line is calculated by using which of the following ratios?

1. R_{source} to R_{load} of the line
2. I_{max} to I_{min} at every point along the line
3. E to I at every point along the line
4. E_{in} to E_{out} of the line

3-9. Maximum transfer of energy from the source to the transmission line takes place when what impedance relationship exists between the source and the transmission line?

1. When the load impedance equals the source impedance
2. When the load impedance is twice the source impedance
3. When the load impedance is half the source impedance
4. When the load impedance is one-fourth the source impedance

3-10. Which of the following sets of terms represents a type of loss in a transmission line?

1. I^2R and induction only
2. Induction and dielectric only
3. Dielectric and radiation only
4. I^2R, induction, and dielectric

3-11. Skin effect is classified as which of the following types of loss?

1. Copper
2. Voltage
3. Induction
4. Dielectric

3-12. What transmission-line loss is caused by magnetic lines of force not returning to the conductor?

1. Copper
2. Radiation
3. Induction
4. Dielectric

3-13. When a dc voltage is applied to a transmission line and the load absorbs all the energy, what is the resulting relationship between current and voltage?

1. They are in phase with each other
2. They are equal to Z_o of the line
3. They are out of phase with each other
4. They are evenly distributed along the line

3-14. The initial waves that travel from the generator to the load of a transmission line are referred to as what type of waves?

1. Incident
2. Refracted
3. Reflected
4. Diffracted

3-15. Waves that travel from the output end to the input end of a transmission line are referred to as what type of waves?

1. Incident
2. Refracted
3. Reflected
4. Diffracted

3-16. The ratio of maximum voltage to minimum voltage on a transmission line is referred to as the

1. rswr
2. pswr
3. vswr
4. iswr

3-17. Which of the following ratios samples the magnetic field along a line?

1. Vswr
2. Pswr
3. Iswr
4. Rswr

3-18. Which of the following lines is NOT a transmission medium?

1. Load line
2. Coaxial line
3. Two-wire open line
4. Twisted-pair line

3-19. Electrical power lines are most often made of which of the following types of transmission lines?

1. Twin-lead line
2. Shielded-pair line
3. Two-wire open line
4. Two-wire ribbon line

3-20. Uniform capacitance throughout the length of the line is an advantage of which of the following transmission lines?

1. Coaxial line
2. Twisted pair
3. Shielded pair
4. Two-wire open line

3-21. What is the primary advantage of a rigid coaxial line?

1. Low radiation losses
2. Inexpensive construction
3. Low high-frequency losses
4. Easy maintenance

3-22. The most efficient transfer of electromagnetic energy can be provided by which of the following mediums?

1. Waveguides
2. Twin-lead flat lines
3. Single-conductor lines
4. Coaxial transmission lines

3-23. Copper I^2R losses are reduced by what physical property of waveguides?

1. Small surface area
2. Large surface area
3. Shape of the waveguides
4. Waveguide material being used

3-24. In a coaxial line, the current-carrying area of the inner conductor is restricted to a small surface layer because of which of the following properties?

1. Skin effect
2. Copper loss
3. Conductor density
4. Waveguide material being used

3-25. Which of the following dielectrics is used in waveguides?

1. Air
2. Mica
3. Insulating oil
4. Insulating foam

3-26. Which of the following characteristics of a waveguide cause its lower-frequency limitation?

1. I^2R loss
2. Physical size
3. Wall thickness
4. Dielectric loss

3-27. At very high frequencies, ordinary insulators in a two-wire transmission line display the characteristics of what electrical component?

1. An inductor
2. A resistor
3. A capacitor
4. A transformer

3-28. At high frequencies, which of the following devices works best as an insulator?

1. An open half-wave section
2. An open quarter-wave section
3. A shorted half-wave section
4. A shorted quarter-wave section

3-29. The range of operating frequencies is determined by which of the following waveguide dimensions?

1. The widest (height/width)
2. The narrowest (height/width)
3. The shortest (length)
4. The longest (length)

3-30. The cutoff frequency for a waveguide is controlled by the physical dimensions of the waveguide and is defined as the frequency at which two quarter wavelengths are

1. shorter than the "a" dimension
2. shorter than the "b" dimension
3. longer than the "a" dimension
4. longer than the "b" dimension

3-31. In practical applications, which of the following dimensions describes the wide dimension of the waveguide at the operating frequency?

1. 0.1 wavelength
2. 0.2 wavelength
3. 0.5 wavelength
4. 0.7 wavelength

3-32. Which of the following fields is/are present in waveguides?

1. E field only
2. H field only
3. E and H fields
4. Stationary fields

3-33. A difference in potential across a dielectric causes which of the following fields to develop?

1. Electric only
2. Magnetic only
3. Electromagnetic

3-34. H lines have which of the following distinctive characteristics?

1. They are continuous straight lines
2. They are generated by voltage
3. They form closed loops
4. They form only in the waveguide

3-35. For an electric field to exist at the surface of a conductor, the field must have what angular relationship to the conductor?

1. 0°
2. 30°
3. 45°
4. 90°

3-36. If the wall of a waveguide is perfectly flat, the angle of reflection is equal to which of the following angles?

1. Cutoff
2. Incidence
3. Refraction
4. Penetration

3-37. The cutoff frequency in a waveguide occurs at exactly what angle of reflection?

1. 10°
2. 30°
3. 45°
4. 90°

3-38. How does the group velocity of an electromagnetic field in a waveguide compare to the velocity of a wavefront through free space?

1. Group velocity is somewhat faster
2. Group velocity is somewhat slower
3. Group velocity is twice that of free velocity
4. Free velocity is twice that of group velocity

3-39. The group velocity of a wavefront in a waveguide may be increased by which of the following actions?

1. Decreasing the frequency of the input energy
2. Increasing the frequency of the input energy
3. Increasing the power of the input energy
4. Decreasing the power of the input energy

3-40. The various field configurations that can exist in a waveguide are referred to as

1. wavefronts
2. modes of operation
3. fields of operation
4. fields of distribution

3-41. The most efficient transfer of energy occurs in a waveguide in what mode?

1. Sine
2. Dominant
3. Transverse
4. Time-phase

3-42. How is the cutoff wavelength for a circular waveguide computed?

1. 1.17 times the radius of the waveguide
2. 1.17 times the diameter of the waveguide
3. 1.71 times the diameter of the waveguide
4. 1.71 times the radius of the waveguide

3-43. The field configuration in waveguides is divided into what two categories?

1. Half-sine and dominant
2. Transverse electric and transverse magnetic
3. Transverse electric and dominant
4. Transverse magnetic and half-sine

3-44. With a mode description of $TE_{1,0}$, what maximum number of half-wave patterns exist across the "a" dimension of a waveguide?

1. One
2. Two
3. Three
4. Four

3-45. With the mode description, $TE_{1,1}$, what maximum number of half-wave patterns exist across the diameter of a circular waveguide?

1. One
2. Two
3. Three
4. Four

3-46. Which of the following devices CANNOT be used to inject or remove energy from a waveguide?

1. A slot
2. A loop
3. A probe
4. A horn

3-47. Loose coupling is a method used to reduce the amount of energy being transferred from a waveguide. How is loose coupling achieved when using a probe?

1. By doubling the size of the probe
2. By increasing the length of the probe
3. By decreasing the length of the probe
4. By placing the probe directly in the center of the energy field

3-48. Increasing the size of the loop wire increases which of the following loop capabilities?

1. Efficiency
2. Bandwidth coverage
3. Power-handling capability
4. Each of the above

3-49. A waveguide that is not perfectly impedance matched to its load is not efficient. Which of the following conditions in a waveguide causes this inefficiency?

1. Sine waves
2. Dominant waves
3. Standing waves
4. Transverse waves

3-50. A waveguide iris that covers part of both the electric and magnetic planes acts as what type of equivalent circuit at the resonant frequency?

1. As an inductive reactance
2. As a shunt resistance
3. As a capacitive reactance
4. As a shorted 1/4 wave stub

3-51. A horn can be used as a waveguide termination device because it provides which of the following electrical functions?

1. A reflective load
2. An absorptive load
3. An abrupt change in impedance
4. A gradual change in impedance

3-52. For a waveguide to be terminated with a resistive load, that load must be matched to which of the following properties of the waveguide?

1. The bandwidth
2. The frequency
3. The inductance
4. The characteristic impedance

3-53. A resistive device with the sole purpose of absorbing all the energy in a waveguide without causing reflections is a/an

1. iris
2. horn
3. antenna
4. dummy load

3-54. A resistive load most often dissipates energy in which of the following forms?

1. Heat
2. Light
3. Magnetic
4. Electrical

3-55. Reflections will be caused by an abrupt change in which of the following waveguide's physical characteristics?

1. Size and shape only
2. Size and dielectric material only
3. Dielectric material and shape only
4. Size, shape, and dielectric material

3-56. A waveguide bend that in the E and H plane must be greater than two wavelengths to prevent

1. cracking
2. reflections
3. energy gaps
4. electrolysis

3-57. A flexible waveguide is used in short sections because of the power-loss disadvantages. What is the cause of this power loss?

1. Walls are not smooth
2. E and H fields are not perpendicular
3. Cannot be terminated in its characteristic impedance
4. Wall size cannot be kept consistent

3-58. The choke joint is used for what purpose in a waveguide?

1. To reduce standing waves
2. To restrict the volume of electron flow
3. To prevent the field from rotating
4. To provide a joint that can be disassembled during maintenance

3-59. A circular waveguide is normally used in a rotating joint because rotating a rectangular waveguide would cause which of the following unwanted conditions?

1. Oscillation
2. Large power loss
3. Decrease in bandwidth
4. Field-pattern distortion

3-60. In your waveguide inspection, you should be alert for which of the following problems?

1. Corrosion
2. Damaged surfaces
3. Improperly sealed joints
4. Each of the above

3-61. What type of corrosion occurs when dissimilar metals are in contact with each other?

1. Contact
2. Metallic
3. Electrical
4. Electrolytic

3-62. Internal arcing in a waveguide is usually a symptom of which of the following conditions?

1. Change in mode
2. Electrolysis at a joint
3. Moisture in the waveguide
4. Gradual change in frequency

3-63. What is the primary purpose of a directional coupler?

1. To sample the energy in a waveguide
2. To change the phase of the energy in the waveguide
3. To change the direction of energy travel in the waveguide
4. To allow energy in the waveguide to travel in one direction only

3-64. What is the electrical distance between the two holes in a simple directional coupler?

1. 1/8 wavelength
2. 1/4 wavelength
3. 1/2 wavelength
4. 3/4 wavelength

3-65. Of the following characteristics, which is NOT required for a device to be considered a resonant cavity?

1. Be enclosed by conducting walls
2. Possess resonant properties
3. Contain oscillating electromagnetic fields
4. Be round or elliptical in shape

3-66. What property gives a resonant cavity a narrow bandpass and allows very accurate tuning?

1. Low Q
2. High Q
3. Inductive reactance
4. Capacitive reactance

3-67. What factor(s) determine(s) the primary frequency of a resonant cavity?

1. Size only
2. Shape only
3. Size and shape
4. Q of the cavity

3-68. Tuning is the process of changing what property of a resonant cavity?

1. The Q
2. The power
3. The cutoff frequency
4. The resonant frequency

3-69. What are the two basic types of waveguide T junctions?

1. H and T
2. H and E
3. Hybrid Ring and magic T
4. Q and magic T

3-70. A waveguide junction in which the arm area extends from the main waveguide in the same direction as the electric field is an example of what type junction?

1. E-type T
2. H-type T
3. H-type T junction
4. H-type junction

3-71. Low power handling capabilities and internal power losses are the primary disadvantages of which of the following junctions?

1. Magic T
2. Rat race
3. Duplexer
4. Hybrid ring

3-72. The hybrid ring is usually used as what type of device in radar systems?

1. Mixer
2. Detector
3. Duplexer
4. Impedance matcher

Reconfigurable Antennas

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Abstract

Reconfigurable antennas change polarization, operating frequency, or far-field pattern in order to cope with changing system parameters. This paper reviews some of the past and current technology applicable to reconfigurable antennas, with several examples of implementations. Mechanically movable parts and arrays are discussed, as well as more-recent semiconductor-component and tunable-material technologies applicable to reconfigurable antennas.

Keywords: Reconfigurable antennas; reconfigurable arrays; antenna arrays; wideband antennas; reflector antennas; microstrip patch antennas

1. Introduction

To configure means to arrange or organize the parts of something to achieve a purpose. For instance, configuring a microstrip antenna consists of determining the patch shape, substrate parameters, type and location of the feed, etc., in order for the antenna to radiate at a desired frequency and polarization. If the desired operating characteristics of the antenna change, then the antenna must be reconfigured or rebuilt to meet the new specifications. Reconfigurable antennas change their performance characteristics by altering the current flow on an antenna, using mechanically movable parts, phase shifters, attenuators, diodes, tunable materials, or active materials. A reconfigurable antenna can be a single antenna or an array.

A reconfigurable antenna modifies the antenna's pattern, polarization, or frequency/bandwidth in some desirable fashion. For instance, a reconfigurable antenna might steer or place a null in an antenna's pattern, switch from right-hand circularly polarized to left-hand circularly polarized, or move the resonant frequency from 2 GHz to 2.4 GHz. In this way, a single antenna replaces two or more antennas, in order to achieve multiple goals. Arrays are the ultimate reconfigurable antenna, because they have many avenues for controlling the current across the aperture. A reconfigurable array may have reconfigurable elements, or a reconfigurable array may change the amplitude and phase of each element, the number of active elements, the elements' positions, or the polarization.

The idea of reconfiguring an antenna is relatively old. In the early 1930s, the nulls of a two-element array were steered by using a calibrated variable phase changer in order to determine the direction of arrival of a signal [1]. Bruce and Beck changed the size of the rhombic antenna in Figure 1 by stretching the wires with a motor and weights [2]. The element lengths of the rhombus were 184 m, and the interior angles were between 132° and 164°. Reconfiguring the antenna consisted of adjusting the distance of the major axis by more than 100 m. Changing the shape of the rhombic wire antenna steered the beam in elevation. The Multiple-Unit Steerable Antenna (MUSA) was a six-element array of rhombic antennas with phase shifters at five of the elements [3]. A linear phase shift steered the main beam. This concept was perfected and introduced for azimuth scanning with a 14-row by three-column array of polyrod antennas. This array had 13 rotary phase changers for beam steering [4]. The Wullenweber array was invented during WW II. It is a large direction-finding circular array with a narrow beam that scans 360° in azimuth by activating a small group of adjacent elements along the circle [5].

In 1979, "reconfigurability" was defined as "the ability to adjust beam shapes upon command" [6]. The authors used

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a six-beam antenna to dynamically change the coverage area for a communications satellite. Several additional papers reported other reconfigurable space-based arrays. In the 1990s, a research group in England described their efforts to alter the reflecting surface of a parabolic-reflector antenna in order to control the radiation pattern [7]. From the mid-1990s until the present, reconfigurable-antenna projects have mostly involved microstrip antennas, and various semiconductor technologies applied to altering the current flow on the microstrip antenna. Bernhard wrote an excellent overview of reconfigurable antennas, with many examples [8].

This paper is an introduction/tutorial on reconfigurable-antenna technology. It begins with the oldest technology, mechanically movable parts and arrays, and then transitions to the more-recent semiconductor-component and tunable-material technologies. A detailed explanation of the technology with background information and examples are given. Some of the interesting material technologies, such as active materials, are promising for reconfigurable antennas with low loss at high frequencies.

2. Mechanically Movable Parts

The first reconfigurable antennas had mechanically movable parts (as in Figure 1). Large reflector antennas moved feeds to steer the main beam, or interchanged feed antennas to change frequency bands. An excellent example is the Arecibo spherical-reflector antenna, shown in Figure 2. A Gregorian dome and a carriage house with linear antennas are on either side of a curved track, called the azimuth arm. Electric motors move the Gregorian dome and carriage house many meters from vertical to 20° off vertical, in addition to rotating 360° in azimuth, in order to steer the main beam. Inside the Gregorian dome, the secondary and tertiary reflectors focus the radiation onto one of several receiving antennas, which can be moved into position using a large rotating platform. Hanging below the carriage house are various linear antennas, each tuned to a narrow band of frequencies. A second example is a beam-waveguide antenna. This type of Cassegrain reflector guides the signal by mirrors, from its normal focal point near the vertex, to a focal point below the structural support (Figure 3) [9]. The rotating reflector at the base of the antenna steers the signal to an appropriate feed. In this way, a single reflector antenna can cover a very broad frequency range.

It is also possible to place nulls in the antenna pattern by altering the reflector's surface. One approach to reconfiguring a reflector's surface is to move scattering plates near the surface to cause cancellation of interfering signals entering the sidelobes [10, 11]. Distorting a mesh-reflector's surface using electrostatic controls is another possible approach [7, 12]. The scattering from a small distorted part of the reflector is sufficient to place a null in a sidelobe.

3. Reconfigurable Arrays

Most reconfigurable arrays control the array pattern in order to steer the beam, place a null, modify the main beam shape, or change the sidelobe levels. There have been several applications that involve changing the polarization or frequency of operation, as well. The textbook approach to reconfiguring an array has adjustable amplitude and/or phase weights at each element. The literature on these arrays is vast, and will not be covered here. A digital beamformer significantly enhances the ability of the array to control the radiation pattern, since all the reconfiguring is done in software.

Beam control is possible by turning elements on and off in the array, as was mentioned earlier with a Wullenweber array.

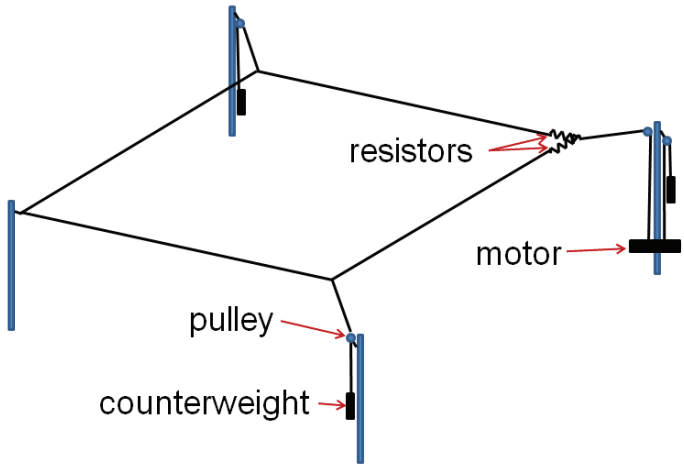

Figure 1. A steerable horizontal rhombic antenna.

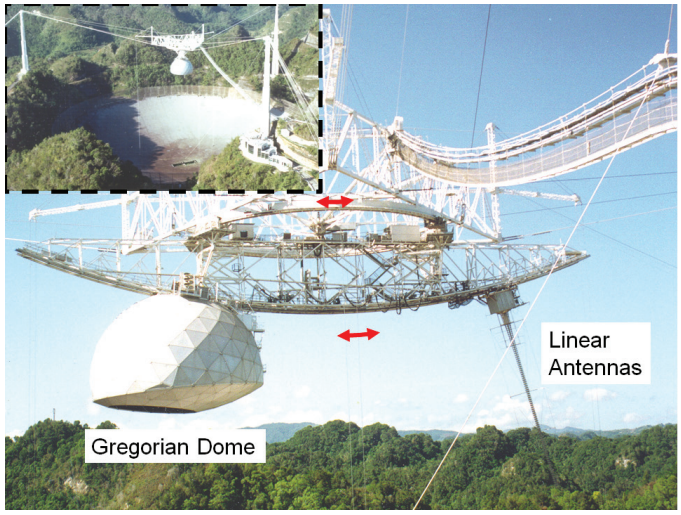

Figure 2. A blown-up picture of the feed for the Arecibo antenna (upper left) (photos courtesy of the NAIC – Arecibo Observatory, a facility of the NSF).

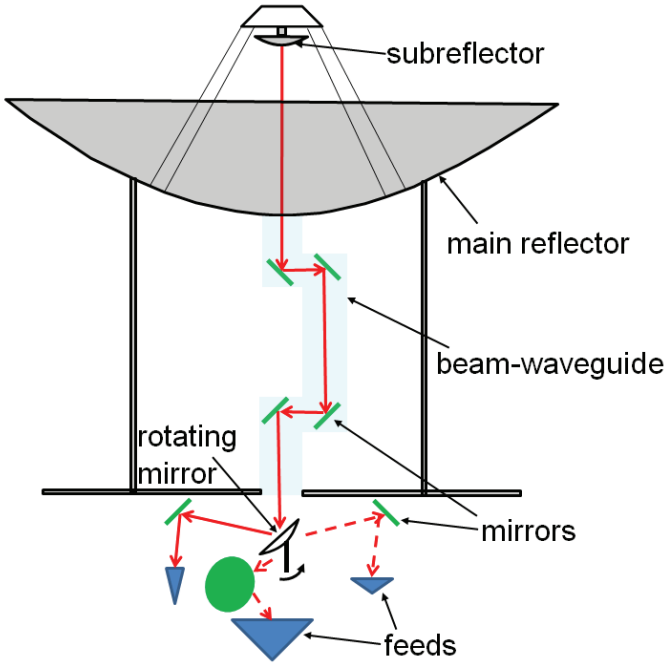

Figure 3. A diagram of a beam-waveguide antenna.

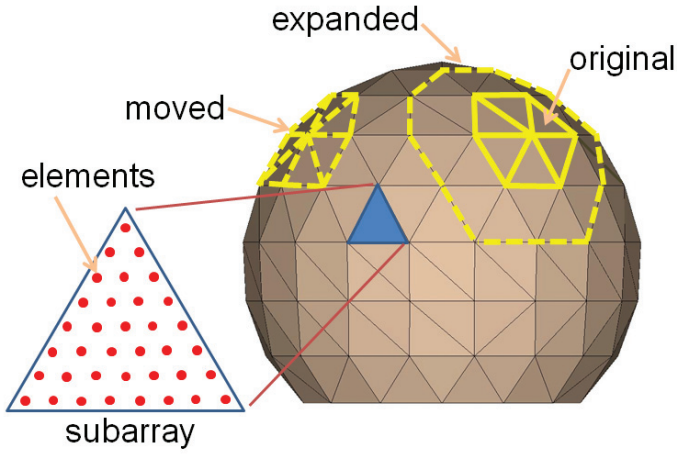

Figure 4. A phased array built on a geodesic dome. The yellow regions represent regions of activated subarrays.

Figure 4 shows this concept extended to a spherical array that was implemented on a geodesic dome. Each triangle on the dome was a subarray [13]. The active aperture was formed by combining the outputs from adjacent subarrays. Including more subarrays expanded the original active aperture (Figure 4). The active aperture could be moved to another location on the dome in order to point the main beam in a different direction. Other reconfigurable arrays that make use of switching an element on and off include dynamically thinning an array as one way to alter the sidelobe levels and nulls in order to reject interference [14], and time-modulated arrays that lower the average sidelobe levels by quickly switching elements on and off, with on times proportional to the desired amplitude of the element weight [15].

Another approach to reconfiguring an array is to physically move elements. Some very large reflector-antenna arrays use this concept. For instance, the Atacama Large Millimeter/sub-millimeter Array, or ALMA, will be a radio telescope composed of 66 12 m and 7 m parabolic dishes in Chile's Andes Mountains (Figure 5) [16]. The telescope will operate at frequencies from 31.25 GHz to 950 GHz. Array configurations from 250 m to 15 km will be possible. The ALMA antennas will be moved between flat concrete slabs by a special vehicle, as shown in Figure 5. A second example is the Very Large Array (VLA) radio telescope in New Mexico, built from 27 dish antennas that are 25 m in diameter [17]. Elements are placed in a Y shape, and can be changed by moving the elements along a railroad track (Figure 6). Four possible configurations, with aperture sizes of 36 km, 10 km, 3.6 km, or 1 km, are used. The array can operate between 73 MHz and 50 GHz.

The idea of moving elements to form a desired array aperture also extends to space. For instance, the TechSat 21 project investigated creating a large sparse antenna array by placing antenna elements on small spacecraft, and assembling the aperture with a formation of the satellites [18]. Beam formation was a function of the satellite constellation and the weighting of the signals.

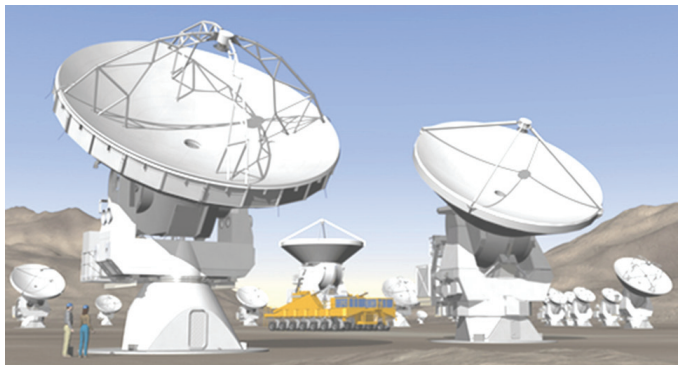

Figure 5. An artist's concept of ALMA (ESO/NAOJ/NRAO).

Figure 6. The VLA antennas on railroad tracks (NRAO/AUI/NSF).

4. RF Switches

An RF switch serves to open or close a current path on a reconfigurable antenna. A popular way to build a reconfigurable antenna is to connect various pieces of the antenna with RF switches. Opening and closing switches guides the current in a desirable path that changes the antenna's radiation properties, as well as its impedance.

RF switches may be mechanical or semiconductor. A switch is an open circuit when no actuation voltage is applied, and a low-impedance path for the RF signal when an actuation voltage is applied. The switch can be implemented in a series or shunt configuration. Some important characteristics of a switch are [19]:

- Characteristic impedance: if the switch is non-reflective, then it is matched to the transmission path; otherwise, it is reflective. The VSWR or return loss indicate the degree of matching.
- Bandwidth: some switches are low-pass filters (pass dc), while others are bandpass filters.
- Topology: is the switch normally on or off?
- Insertion loss and isolation: the ratio between the switch's output and input powers when the switch is on and off, respectively.
- Switching speed: the on time for a switch is defined by the time from when the control pulse reaches 50% of its level to the time that the RF signal is at 90% of its peak. The off time for a switch is defined by the time from when the control pulse reaches 50% of its level to the time that the RF signal is at 10% of its peak. Switching time is the larger of the on and off switch times.
- Expected life time: measured in number of switch actuations until failure.
- Power handling: measured in watts.

Below microwave frequencies, a low-pass switch acts like a resistor (R_{on}) when on, and a capacitor (C_{off}) when off. In contrast, a bandpass switch behaves like one capacitor when on, and a different capacitor when off. As the frequency increases, the ground inductance, bond-wire inductance, and transmission-line properties become significant factors, and complicate the circuit model. Parasitic resistances in the switch limit the upper-frequency bound. A switch figure-of-merit (FOM) is the cutoff frequency [20]:

$$f_c = \frac{1}{2\pi C_{off} R_{on}}. \qquad (1)$$

A rule of thumb for the highest operating frequency for the switch is approximately $f_c/10$ [21].

The switching speeds of semiconductor switches are of the order of nanoseconds, while the switching speeds of mechanical switches are of the order of milliseconds. System constraints may limit the type of switch used, based on required switching times.

Switches can be used to reconfigure microstrip-patch antennas in several different ways. For instance, Figure 7 shows a two-dimensional array of metal patches on a substrate [22]. The sides of adjacent patches are connected with RF switches, in order to configure a desirable patch antenna. Figure 8 is an example of a patch with slots that can be shorted, in order to switch between polarizations [23].

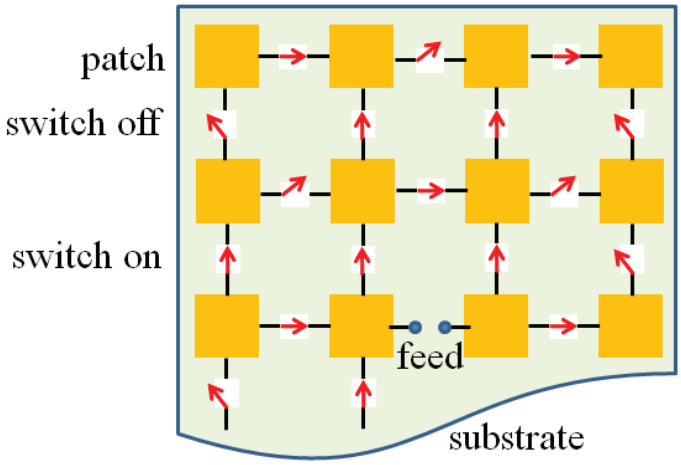

Figure 7. Reconfigurable microstrip patches.

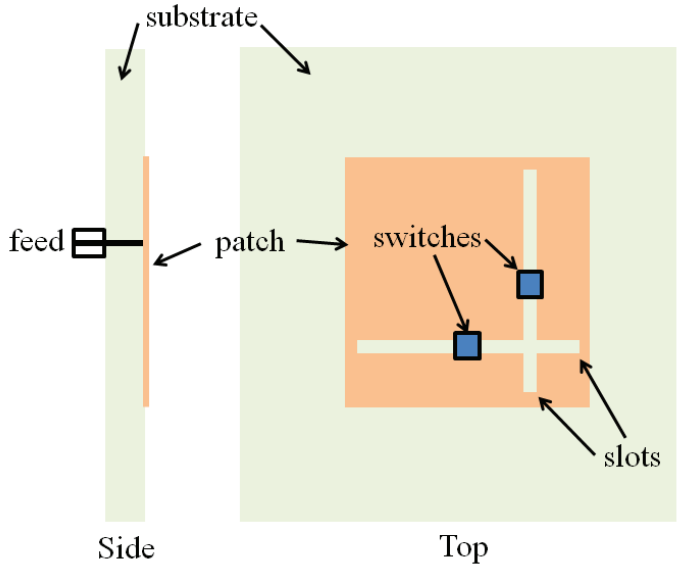

Figure 8. A reconfigurable slotted-patch antenna.

4.1 Semiconductor Switch

Figure 9 is a diagram of a field-effect transistor (FET) switch. Increasing the voltage at the gate increases the conducting channel's size beneath the gate, and allows current to flow between the source and drain. FET switches come in several varieties. Metal-semiconductor field-effect transistor (MESFET) and pseudomorphic high-electron-mobility transistor (PHEMT) switches are two variations, which are compared in Table 1. The authors in [24] suggested using FET switches in a configuration similar to Figure 7. The RF switch array was controlled with a light-emitting-diode backplane, which isolated the control circuitry from the RF paths in the antenna.

Another widely used microwave switch is the PIN diode [25]. It has heavily doped p-type and n-type regions (used for ohmic contacts), which are separated by a wide, lightly-doped intrinsic region (Figure 10). Forward biasing a PIN diode creates a very low resistance at high frequencies, while reverse biasing results in an open circuit. The PIN diode is current controlled, and can handle one or more amps of RF current. The authors in [23] suggested using PIN-diode switches for the configuration in Figure 8.

Table 1 compares the characteristics of FET and PIN-diode switches. Some notable differences between the two types of switches are [26, 27]:

- PIN diodes are current controlled, while FETs are voltage controlled. FETs require only a voltage signal for switching, instead of a dc current. This means that they have essentially zero dc power consumption, compared to the 10 mW or so it takes to turn on a PIN diode.
- PIN diodes have the ability to control large RF signal power while using much smaller levels of control power.
- PIN diodes are less susceptible to electrostatic-discharge (ESD) damage.
- The off-capacitance of PIN diodes is a function of reverse voltage: the more negative the voltage, the less capacitance.
- For FETs, the capacitance is not a strong function of reverse voltage.
- PIN diodes have a higher figure-of-merit than FETs. The useful upper-frequency-response limit of PIN diodes can be much higher, due to lower off-state capacitance (C_{off}) for a given on resistance (R_{on}).

Recently, a reconfigurable patch antenna was demonstrated in which the resonant frequency was altered between 2 GHz and 2.25 GHz, when the electrical length of a spiral-shaped slot in the patch was modified by a biased diode [28].

4.2 MEMS Switches

MEMS switches are tiny mechanical switches made on a substrate (silicon, quartz, glass) [29]. Unlike the PIN-diode and FET switches, a MEMS switch is mechanical. Figure 11 shows three types of MEMS switches in their on and off positions. The cantilever beam in Figures 11a and 11b is anchored to a post on the left, while the other end of the beam is suspended above the drain. An electrostatic force pulls the beam

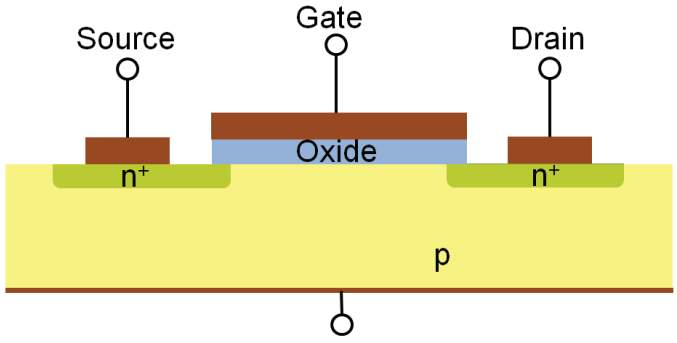

Figure 9. A diagram of a FET.

Table 1. A comparison of FET and PIN-diode switches [21].

	1 µm MESFET	0.25 µm PHEMT	Silicon PIN diode
Number of terminals	3	3	2
Typical on resistance	1.5 ohm-mm	1.2 ohm-mm	1.7 ohms
Typical off capacitance	0.40 pF/mm	0.32 pF/mm	0.05 pF
FOM (Eq. 1)	265 GHz	414 GHz	1872 GHz
Breakdown voltage	15 V	8 V	50 V
Lower frequency limit	dc	dc	10 MHz
Driver circuit complexity	low	low	high
Driver requirements	0 V on; –5 V off	+0.5 V on; –5 V off	5 to 10 mA on; 0 to –30 V off

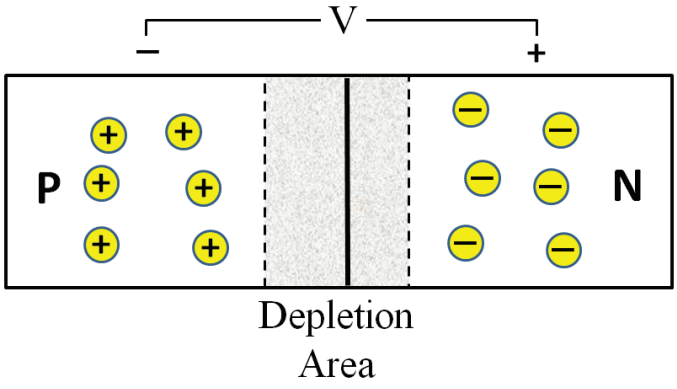

Figure 10. A diagram of a PIN diode.

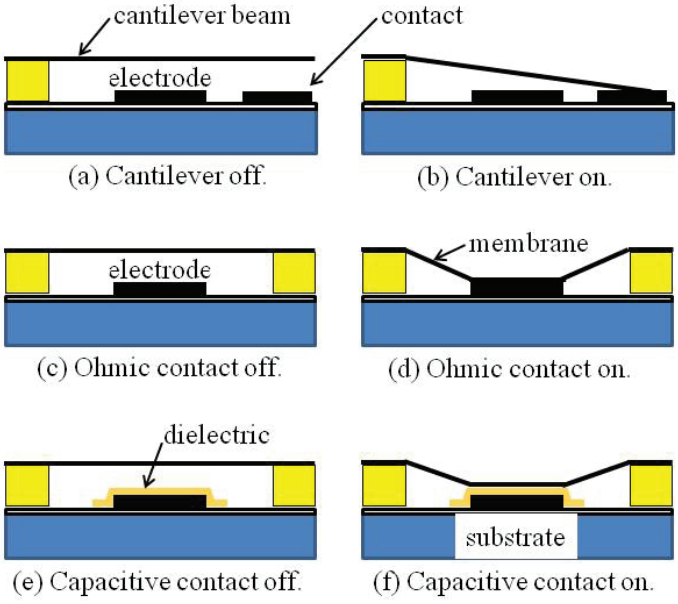

Figure 11. Three types of MEMS switches in the on and off positions.

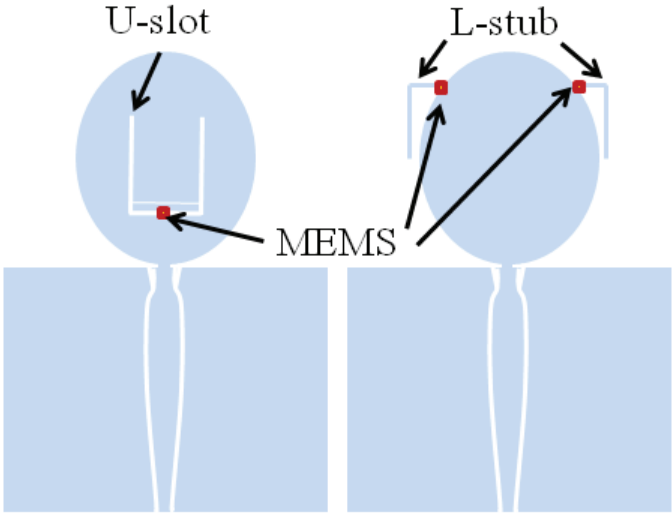

Figure 12. A diagram of the reconfigurable U-slot and L-stub antennas.

down when a voltage is applied, and creates an electrical path between the beam and the drain. Figures 11c-11f show a MEMS membrane switch, which consists of a flexible, thin, metal membrane, anchored to posts at both ends. A potential applied to the bias electrode pulls the membrane down and closes the circuit. An ohmic contact is a metal-to-metal connection, while a capacitive contact has a dielectric between the two metal contacts. Ohmic switches have a higher bandwidth than capacitive switches. The capacitive MEMS switches handle higher peak power levels and tolerate environmental effects better than ohmic contacts. The figure-of-merit for MEMS switches was reported at 9000 GHz [20]. Switching speeds for electro-statically driven capacitor structures are 10 μs. Recently, MEMS switches with piezoelectric films have been developed, and have fast switching times (1-2 μs). The cantilever device travels 6 μm between the on and off states [30].

MEMS switches have low power consumption, low insertion loss, and high isolation, like mechanical switches, but are small, light weight, and low cost, like semiconductor switches [31]. On the other hand, MEMS switches have high losses at microwave and mm-wave frequencies, limited power-handling capability (~100 mW), and they may need expensive packaging to protect the movable MEMS bridges against the environment.

MEMS switches have been proposed for use in reconfigurable antennas since the late 1990s [29]. For example, two ultra-wideband monopoles with a reconfigurable band notch in the wireless-LAN frequency range (5.150 GHz to 5.825 GHz) are shown in Figure 12 [32]. The antenna was an elliptical patch, fed with a coplanar-waveguide line. The U-shaped slot was approximately $\lambda/2$ long, and had a frequency notch when the MEMS switch was open, but not when the MEMS switch was closed. When the switch was open at 5.8 GHz, the currents in the inner and outer side of the slot flowed in opposite directions and cancelled each other. When the MEMS switch was closed, the slot was shorted at its center point, so the total length of the slot was cut in half. Consequently, the slot no longer supported the resonating currents, and radiation occurred as if the slot was not present.

The second antenna in Figure 12 had two inverted-L-shaped open stubs that connected and disconnected the stubs from the elliptical patch via MEMS switches. Shorting the stubs to the patch created a rejection band. At the resonance frequency, the direction of the current on the stub flowed in the opposite direction to the current along the nearby edge of the radiator, so the radiated fields cancelled. When the stubs were not shorted, the antenna operated over the whole UWB range (3.1 GHz to 10.6 GHz) without any rejection band. The MEMS switches actuated through the RF signal path, without any dc bias lines that might have complicated the switch's fabrication and integration, while degrading the radiation performance due to RF leakage through the bias.

Reconfigurable antennas can also change the antenna's polarization. For instance, placing MEMS switches in the feeds of microstrip antennas provides the ability to switch from one linear polarization to the orthogonal linear polarization, or to circular polarization, as shown in Figure 13 [33].

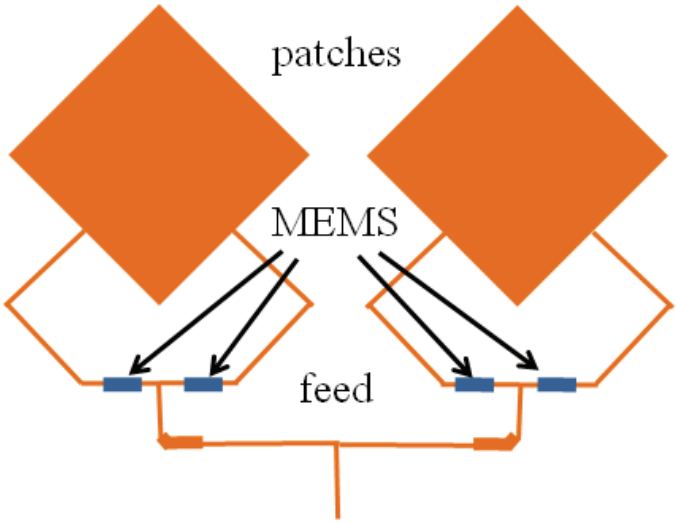

Figure 13. A reconfigurable antenna that changes polarization using MEMS.

MEMS structures are fabricated using a number of techniques, including etching trenches in the substrate, or building a multilayer structure on a planar substrate. Future research and development activities will need to address the cost and scalability of the MEMS process for large arrays. There are several important substrate properties that must be controlled, including surface roughness, permittivity, and dielectric loss. Recently, low-cost liquid-polymer substrates with $\varepsilon_r = 3$ and $\tan\delta = 0.002$ have been shown to be viable substrates for MEMS devices [34].

5. Varactors

Unlike the PIN diode, a varactor diode has a very thin depletion layer that acts like the insulating dielectric, and the P and N regions that act like the conducting plates (Figure 10). The capacitance is inversely proportional to the square root of the applied voltage, since the thickness of the depletion layer increases with the reverse bias. Varactors have a high-to-low capacitance ratio that is typically six over a voltage change of 0 V to 12 V.

Varactors are useful for tuning the antenna's frequency range. For instance, varactors were placed at the radiating edges of a microstrip-patch antenna in order to increase its very narrow bandwidth to a bandwidth of about 30% [35]. Strategically placing varactors along a slot antenna produced dual bands, with controllable first and second resonant frequencies [36]. A reconfigurable partially-reflecting-surface (PRS) antenna was built. It had a measured realized gain that varied from 10 dBi to 16.4 dBi over the tuning range of 5.2 GHz to 5.95 GHz, as the bias voltage was tuned from 6.49 V to 18.5 V [37]. The antenna had a partially-reflecting-surface material on the top surface, and a phase-agile surface below a microstrip patch. One phase-agile reflection cell was a microstrip-patch resonator, with varactor diodes that varied the resonant frequency of the patch antenna's reflection phase. The 13.5% tuning range extended the operating-frequency range by six times. The antenna's tuning bandwidth was limited by the varactor diode's tuning range for the phase-agile cells, and the antenna's efficiency was limited by the varactor diode's series resistance. A slotted-monopole antenna, modified with a varactor, could be tuned between 1.6 GHz and 2.3 GHz without a significant change in the antenna's efficiency [38]. Radiation pattern control has also been demonstrated using varactors [39].

6. Tunable Materials

Prospective devices with performance rivaling varactors will be based on materials that have tunable electrical, magnetic, and mechanical properties. For example, the propagation wavelength within the RF structure or impedance will be altered by permittivity (ε_r) or permeability (μ_r) changes, and the ability to tune the structure will be fundamentally related to the magnitudes of ε_r and μ_r [40]. Tunable mechanical strain alters either the effective length of the antenna, or modifies the impedance through capacitive coupling. Many concepts, such as tunable permeability, have been previously studied for microwave and millimeter-wave phase shifters [41]. Future tunable materials and devices will advance antenna technology through miniaturization and integration. A representative list of potential technologies for reconfigurable antennas is summarized in Table 2.

6.1 Tunable Conductivity

Conductivity in semiconductors can be tuned through changes in temperature, bias, or light. The conductivity of a semiconductor varies depending on the bandgap energy and the level of doping, and depending on defects in the material. Shining light with a photon energy greater than the bandgap on an intrinsic semiconductor increases the free-electron and hole-charge carriers, which in turn increases the semiconductor's conductivity. Solar cells are a good example.

Applying an electric field to a polymer electrolyte-silver-polyaniline composite alters the material from a state of low to high conductivity. Polymer composites with controllable resistance at microwave frequencies have stable, reproducible switching over more than 1000 test cycles [49].

Phase-change chalcogenides, such as $Ge_2Sb_2Te_5$ (proposed for non-volatile memory applications), also have tunable conductivity. Measurements have demonstrated that amorphous chalcogenides maintain low conductivity values into the GHz frequency range [50]. These become conducting upon recrystallization of the amorphous material. Resistivity changes of nearly three orders of magnitude are possible in small structures, by methods involving laser and Joule heating [51].

Table 2. Potential technologies with tunable dielectric, magnetic, and strain response for frequency-agile devices.

Method	Tunability (%)	Q	Stimulus	Frequency
Dielectric Thin Film [42]	30	45	Electric Field $E = 70$ kV/cm	1-20 GHz
Dielectric Bulk Ceramic [43]	16	>100	Electric Field $E = 15$ kV/cm	1-10 GHz
Magnetic Film [44]	15	$5 < Q < 11$	Magnetic Field 80 kA/m (1000 Oe)	0.3-5 GHz
Magnetic Bulk Ceramic [45]	3.4* (12)	<1700	Magnetic Field	7 GHz
Displacement Microstrip [46]	25* (625)	$40 < Q < 100$	Dielectric Translation (100 μm)	3-7 GHz
MEMS variable capacitor [47]	80	$100 < Q < 300$	Mechanical Displacement 1 μm	0.5-4 GHz
Varactor [48]	60	$30 < Q < 60$	Bias Voltage (20 V)	10 GHz

*Tunable frequency is proportional to the square root of the tunable permeability or permittivity, and the squares of the values are shown for comparison with the other methods (shown in parenthesis).

Silicon is a viable candidate for optically controlling conductivity. In [52], the authors reported on creating planar antennas on a silicon wafer by precisely forming plasma channels in the shape of the desired antenna. Varying the conductivity of silicon using an IR LED was reported for a reconfigurable patch antenna [53], and for a partially adaptive array with broadband monopoles [54].

6.2 Tunable Permittivity

A number of approaches have been explored to achieve economical solutions for a high ability to tune with low losses and fast response, for the purpose of controlling microwave filters and phased arrays. The relative "tunability" of the permittivity of a system is defined as

$$n_r = \frac{\varepsilon_r(0) - \varepsilon_r(E)}{\varepsilon_r(0)}, \qquad (2)$$

where $\varepsilon_r(0)$ and $\varepsilon_r(E)$ are the small-signal relative permittivities without bias and with a bias of strength E [V/cm], respectively. Tunabilities as high as 75% have been reported for $(Ba,Sr)TiO_3$ thin films at 1500 kV/cm [55]. A high electric field is required to generate a large tunability; however, the tuning voltage is low (< 25 V) because the films are in the nanometer-thickness range. A figure-of-merit for tunable dielectric devices has been established, which includes the device loss [56]:

$$K = \frac{Q(0)Q(V_{max})(n_r - 1)^2}{n_r}, \qquad (3)$$

where Q is the inverse of the device's loss; $Q(0)$ and $Q(V_{max})$ are at zero bias and at the maximum voltage, respectively; and n_r is the tunability from Equation (2). K factors as high as 500 have been reported for paraelectric $SrTiO_3$ thin films on $SrRuO_3$ conductors.

Dielectric materials with the highest tunability have paraelectric/ferroelectric transition temperatures that are below the operating temperature. The magnitude of the tunability is directly proportional to the dielectric permittivity, which approaches 1,000 for materials with the highest tunability [58]. The intrinsic dielectric loss increases with frequency, which may be a concern as the operating frequency increases into the mm-wave range [59].

The dielectric permittivities of thin films and bulk materials are altered through an externally applied electric field. Large changes in permittivity under a dc voltage bias have been demonstrated in epitaxial paraelectric $Ba_{0.5}Sr_{0.5}TiO_3$ ("BST") thin films that are deposited on single-crystal $LaAlO_3$ and MgO substrates [42]. The loss of the dielectric film is represented as an average Q value between the zero-field and high-electric-field bias. Bulk BST ceramics offer high tunability, and have higher Q values that BST films. Recent results have shown that high electric fields can be applied to bulk ceramics with nanometer-sized grains, corresponding to higher high overall tunability [43].

6.3 Tunable Permeability

Similar to the tunable response in dielectric materials, the magnetic permeability decreases with the application of a

static field. An early example of a 1.4 cm by 1.8 cm rectangular microstrip patch was printed on a 1.27 mm thick ferrite substrate (Trans-Tech G-113 YIG), and was fed with a coaxial probe near the edge of a long dimension [60]. The center design frequency was 4.6 GHz, which was tunable over a 40% bandwidth when it was magnetically biased in the plane of the substrate, and perpendicular to the resonant dimension. The polarization response of patch antennas on garnet single crystals has also been studied at 5 GHz for externally applied magnetic fields of 600 G. One advantage of higher permeability over high permittivity is that the effective patch dimensions can be reduced, without sacrificing bandwidth [61]. There are fundamental materials limitations for tunable magnetic materials, because the permeability and the highest operating frequency are inversely related [62]. Ferrites are a class of high-frequency magnetic materials with low eddy current and hysteretic losses [63].

Tunable magnetic structures have thin-film and bulk-ceramic forms. Conformal Permalloy films (120 nm thick) were deposited on CMOS structures, having lower conduction and hysteresis losses than bulk NiFe in the GHz frequency range [44]. In addition, eddy-current losses at high frequency were minimized through the inductor's design. Judicious patterning of the magnetic film raises the ferromagnetic resonance, and hence the operating frequency, of the tunable thin-film inductors; however, the magnetic-film technology will be limited to frequencies below 10 GHz, because of the limitation in the ferromagnetic resonance frequency. There are fundamental material limitations for tunable magnetic materials, because the magnitude of μ_r and the highest operating frequency are inversely related [62]. The magnetic response of ferrites is limited to below the K band [64]. Bulk-ceramic ferrite plugs were placed within the center of a dielectric resonator operating in the $TE01\delta$ mode [45]. The resonator was placed on top of a coil, which was used to apply the dc magnetic field. A high Q value was reported because of the low-loss dielectric resonator within the tunable structure.

6.4 Tunable Mechanical

Tunable mechanical materials are also known as active materials, and they change shape when exposed to an electric, magnetic, or thermal stimulus. Electrostrictive materials strain under an applied electric field. Magnetostrictive materials change shape due to an applied magnetic field. A magnetic field applied to a ferromagnetic material aligns the magnetic domains, and increases the length of the material in the direction of the magnetic field.

Shape-memory materials change shape due to a temperature change. Both alloys and polymers have been shown to exhibit this change. Shape-memory materials return to their original shape or size when deformed and subjected to an appropriate stimulus. The first shape-memory alloy, Nitinol (*Ni*ckel *Ti*tanium *N*aval *O*rdance *L*aboratory), was discovered in the early 1960s [65]. Today, copper-aluminum-nickel, copper-zinc-aluminum, nickel-titanium, and iron-manganese-silicon are other commonly used examples of alloys that when plastically deformed at a relatively low temperature, return to their original shape when subjected to a high temperature. Some shape-memory alloys, like Ni-Mn-Ga, respond to magnetic fields [66]. Shape-memory polymers (SMPs) are the polymer equivalent of shape-memory alloys. Light-induced shape-memory polymers are also available [67]. Lu and Kota outlined the use of shape-memory materials to adaptively deform reflector antennas from a parabolic to a spherical shape, or to tilt the reflector's surface [68].

The strain response of active materials is summarized in Table 3. Piezoelectric, electrostrictive, and Maxwell stress-based actuations are mechanical responses to an applied electric field. Piezoelectricity is generally limited to non-centrosymmetric crystal structures, and $Pb(Zr,Ti)O_3$ ("PZT") ceramics are commonly used for actuator applications. Exceptionally high piezoelectric coefficients have been discovered for $Pb(Mg,Nb)O_3$-$PbTiO_3$ ("PMN-PT") single crystals, which have strain values exceeding 1%. Electrostriction is possible in all materials, and the highest strains are found in PMN-PT ceramics and ferroelectric poly(vinylidene fluoride-trifluoroethylene) polymers. Maxwell stresses, created by the electrostatic attraction between oppositely charged conductor plates, cause very high strains in compliant polymers, such as silicone.

Only the strain parameters are shown in Table 2, and additional requirements must be considered for implementation of active materials into a tunable device. For example, the elastic modulus of high-strain polymers is low, and will limit the total force available to move a tunable structure. Piezoelectric materials have hysteretic strain behavior at high electric

Table 3. The strain responses of several types of active materials.

Actuation Mechanism	Piezoelectric	Electrostriction	Maxwell	Magnetostriction	Phase Change
Materials	PZT, PMN-PT single crystal	PMN-PT ceramics and PVDF	Silicone	Terfernol	Nitinol
% strain range	0.1 to 1.5	0.1 to 4	20	0.2	6
Reference	[69]	[70]	[71]	[72]	[72]

fields, which may limit reproducibility in the strain response. The time response and reversibility of the phase transition in Nitinol will limit its use for fast tuning systems.

Tunable micro-electromechanical system (MEMS) and microstrip structures take advantage of piezoelectric or electrostatic forces to move capacitor plates. Tunable microstrip configurations, based on translatable superstrates above ring resonators, offer high tunability and Q values [46]. A variable MEMS capacitor has been demonstrated. The capacitance changed from 1.2 pF to 0.25 pF as the voltage was altered from 15 V to 10 V. The Q factor was high, because the structure was fabricated from high-resistivity silicon [47].

There are several challenges in comparing tunable devices that have a range of impedance, Q, and tunability parameters. Figures-of-merit have incorporated all of these parameters and have been tailored to specific devices, such as phase shifters [40]. In addition, device-performance metrics, such as temperature stability, RF power, response time, and cost, must also be considered [48]. In Table 1, the device tunability is defined by different parameters, and depends on the physical mechanism. For example, the MEMS, dielectric devices, and varactors depend upon a permittivity or capacitance change. The magnetic devices depend upon a permeability or inductance change. The magnetic bulk-ceramic and displacement-microstrip methods that are listed in Table 3 are defined by tunable frequency: hence, these values are squared, to provide a meaningful comparison with the other methods. From Table 3, it is apparent that devices that are based on mechanical strain are promising candidates for tunable devices with high Q and high tunability. In addition, these devices are not limited in frequency, as in the case of tunable magnetic materials.

7. Conclusions

Reconfigurable antennas have been around since the 1930s. Initially, they were based on mechanical movement of a feed or other antenna part. Antenna arrays took reconfigurability to a new level with the electronic control of the antenna's pattern. Semiconductor and MEMS switches have been at the heart of most reconfigurable-antenna research since the late 1990s. More recently, materials research has introduced new approaches to reconfigurable antennas, via tunable constitutive parameters and even active materials that move in response to an electrical stimulus. The need for a single antenna to perform multiple missions will continue the drive for antennas that can reconfigure themselves based upon the current need.

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Introducing the Feature Article Authors

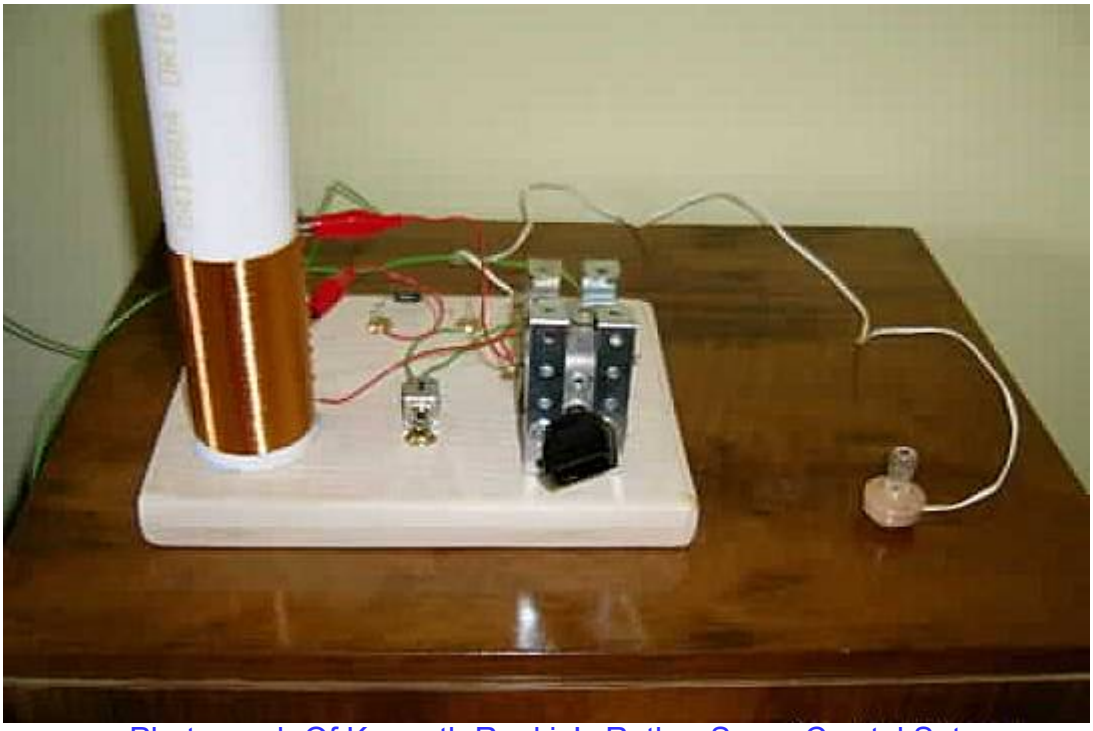
Randy L. Haupt received the BSEE from the USAF Academy (1978), the MS in Engineering Management from Western New England College (1982), the MSEE from Northeastern University (1983), and the PhD in Electrical Engineering from the University of Michigan (1987). He is a Professor and Department Head for the Electrical Engineering and Computer Science Department of the Colorado School of Mines, Golden, Colorado. He was an RF Staff Consultant at Ball Aerospace & Technologies, Corp. He was Senior Scientist and Department Head at the Applied Research Laboratory of Penn State, Professor and Department Head of ECE at Utah State, Professor and Chair of EE at the University of Nevada Reno, and Professor of EE at the USAF Academy. He was a project engineer for the OTH-B radar, and a research antenna engineer for Rome Air Development Center early in his career. He is coauthor of the books *Practical Genetic Algorithms, Second Edition* (Wiley, 2004); *Genetic Algorithms in Electromagnetics* (Wiley, 2007), and *Introduction to Adaptive Antennas* (SciTech, 2010), as well as author of *Antenna Arrays: A Computation Approach* (Wiley, 2010). Dr. Haupt was the Federal Engineer of the Year in 1993. He is a Fellow of the IEEE and of the Applied Computational Electromagnetics Society (ACES). He serves as Associate Editor for the "Ethically Speaking" column in the *IEEE Antennas and Propagation Magazine*.

Michael Lanagan received his BS in Ceramic Engineering from the University of Illinois in 1982, and his PhD from Penn State in 1987. He is presently Professor of Engineering Science and Mechanics. Before joining Penn State, Dr. Lanagan was at Argonne National Laboratory for twelve years. Prof. Lanagan's research interests are in two general areas. The first is the exploration of new materials and structures for energy-storage applications. The second research topic involves advanced materials and devices for enhanced magnetic-resonance imaging. He is an author of over 250 technical publications and 10 patents. His present interests also include synthesis, processing, and characterization of new materials for wireless communications and power electronics applications.

CRYSTAL SETS 4
A Crystal Set By Kenneth Rankin

CRYSTAL SETS 4: A CRYSTAL SET BUILT BY KENNETH RANKIN

MDS975.co.uk reader Roddie Rankin kindly sent in the details of a crystal set built by his son Kenneth for a school project which won him great praise and merit. The crystal set is so nicely made and the project so well written that I thought it deserved a special page of its own.



Photograph Of Kenneth Rankin's Rather Super Crystal Set

CONTENTS

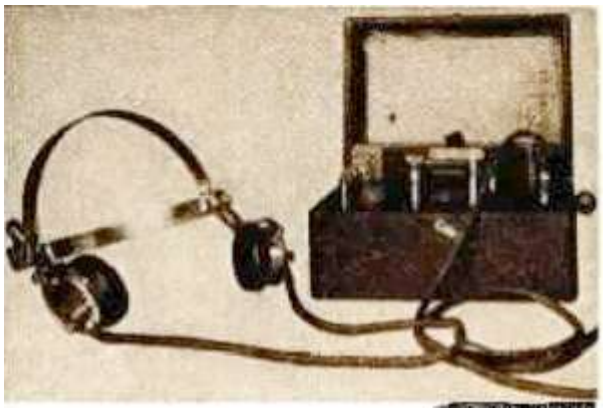
1 - What Is A Crystal set?
2 - How A Crystal Set Works
3 - History Of Crystal Sets
4 - How I Built My Crystal Set
5 - The Results I Obtained
6 - Where I Found The Information
7 - Quiz

1 - WHAT IS A CRYSTAL SET?

Perhaps you think, like I did, that a crystal set is just a collection of lovely stones. In fact a crystal set is a very early, simple domestic radio. The photograph above shows the crystal set I built.

The crystal set does not have a battery. It runs completely from the radio waves it picks up in the air. It is absolutely free to run!

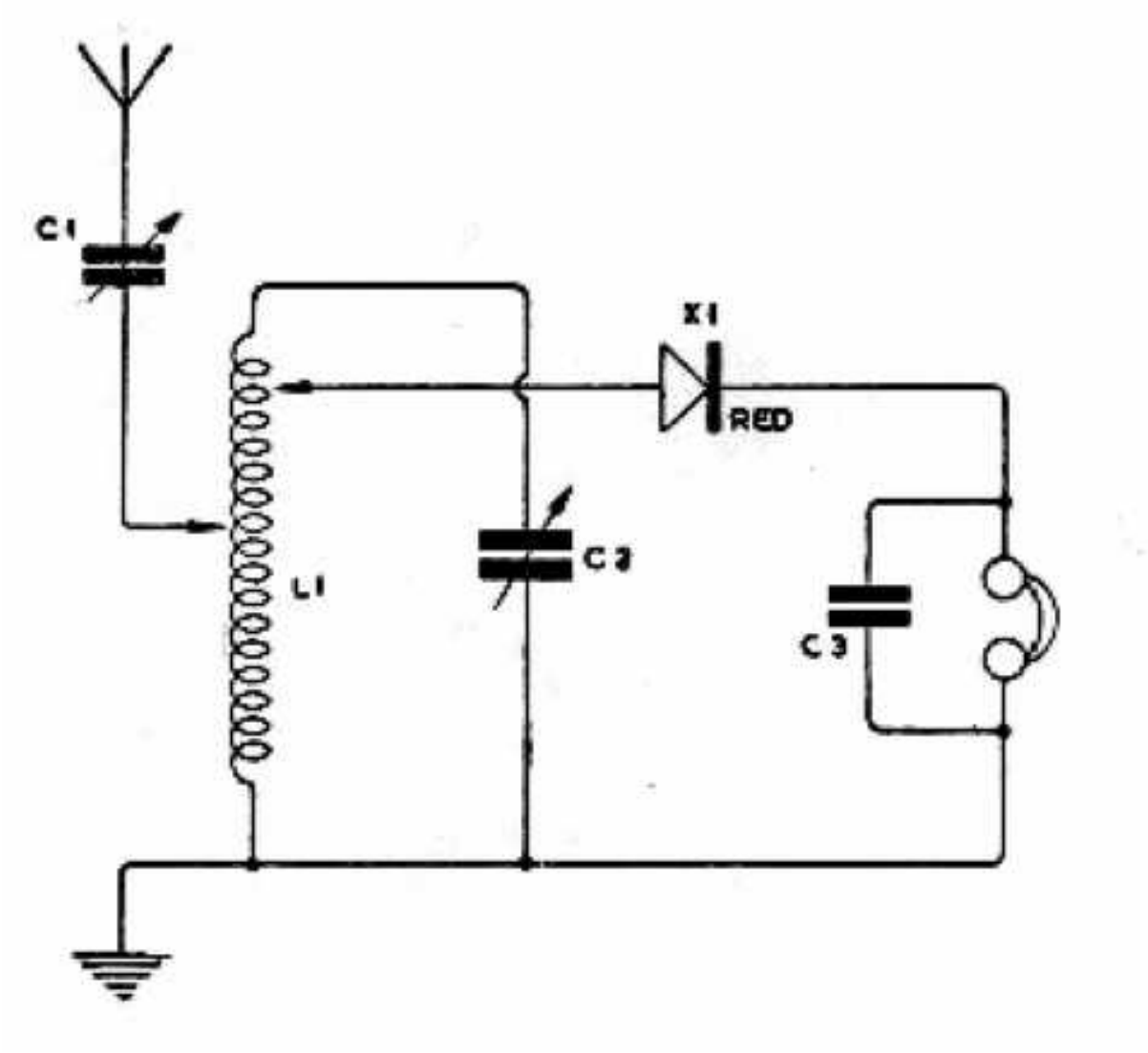
Please read on to find out more about this exciting invention!



The birth of domestic radio: A deluxe crystal set produced by Marconi for the reception of programmes from the first of the BBC's radio stations, 2LO in London.
[from Pageant Of The Century - Odhams Press]

2 - HOW A CRYSTAL SET WORKS

The diagram below shows the circuit diagram of the crystal set I have built. It consists of a coil (L1), a tuning capacitor (C2), a diode (X1), an aerial, a crystal earpiece and an earth. (Capacitors C1 and C3 are not essential.)



Circuit Diagram Of The Crystal Set

First the aerial picks up radio waves from radio stations all over the world. Then the required station needs to be selected, this is called tuning, and this is achieved with a Tuned Circuit. The Tuned Circuit in the crystal set consists of two items; a component called a variable capacitor and a coil of wire.

By adjusting the value of the capacitor and coil tuning is achieved. The coil taking unwanted signals of low frequency straight through to the earth, which gets rid of them, while the capacitor takes unwanted signals of high frequency straight through to the earth. The remaining signal, in between the high frequency and low frequency signals, is the signal of the required station which is not passed to earth by either the coil or the capacitor. This is the frequency to which the set is now tuned in.

You normally change the frequency to which a set is tuned in to by turning the knob on the tuning capacitor, thus receiving different stations by varying the capacitance of the variable capacitor. The number of turns on the coil can also be changed [called varying the inductance of the coil] and this will give different tuning ranges. A larger number of turns provided longer wavelengths [lower frequencies], while a smaller number of turns of wire provides shorter wavelengths [higher frequencies].

[It is also worth noting that the inductance of a coil can be increased by inserting a component called a Ferrite Rod into the centre of the coil.]

Once tuned into the required station this remaining signal goes to earth by travelling through the diode and earpiece. The diode takes the sound signal out of the radio signal [called demodulation], and passes it to the earpiece. The sound signal travels through the crystal in the earpiece which vibrates the air in the earpiece giving us the sound of the radio station to which you are tuned.

3 - THE HISTORY OF THE CRYSTAL SET

The history of crystal sets begins with the first broadcast stations being opened. This happened in England in 1922. The simplest form of receiver for these broadcasts was the crystal set.

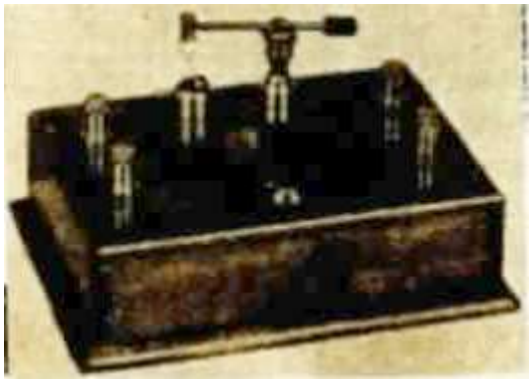
In those early sets they had "cat's whiskers" instead of diodes. A cat's whisker was a fine piece of wire which was adjusted to make a suitable contact with a crystal. The crystal would be made of a substance such as coke or galena. An example from 1923 is shown in the picture below. Notice how well made these instruments were.

There were several limitations to the crystal set: it needed a big aerial, an earth connection, the clumsy cat's whisker, and the weak signal could only be listened to by one person at a time with headphones. Very quickly the crystal set began to be replaced by valve radios with loudspeakers, powered by battery (see the picture from 1923).

In World War II, crystal sets were used by prisoners of war in prison camps to listen to news from home. As they were often badly treated it was hard to get all the right parts to build the set.

By this time crystal sets were no longer used in the home. Instead radios like the attractive 1937 model in the picture were being used.

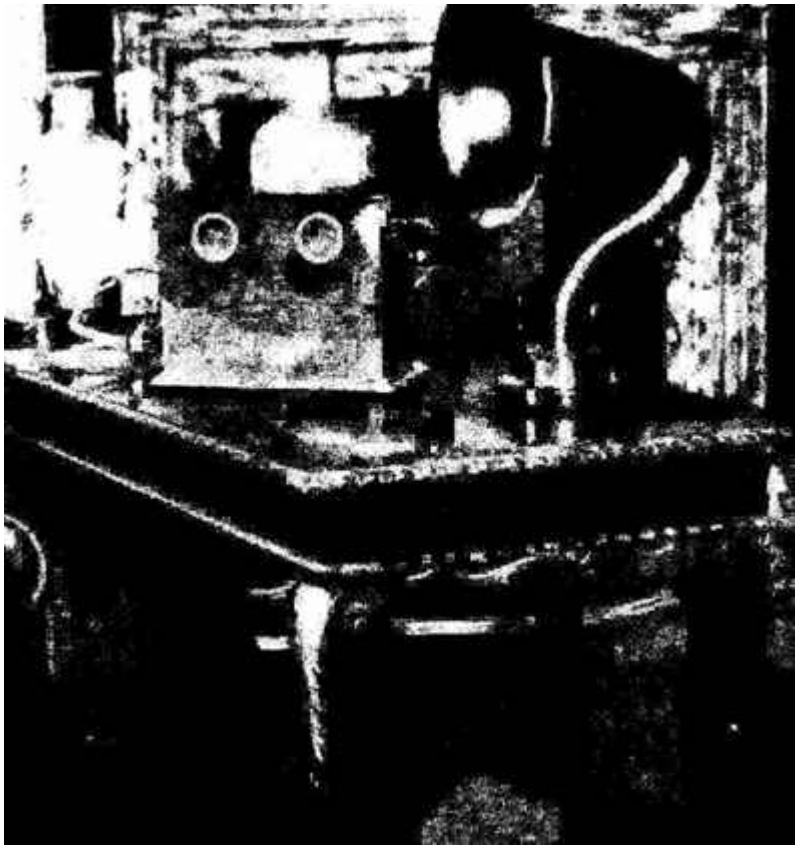
Nowadays people only use crystal sets as a hobby. There are various different kits available on the market (see websites in the references below).



Photograph Of A Crystal Set From 1923

[from Pageant Of The Century - Odhams Press]

Crystal sets were primitive instruments used for 'listening in', and were fitted with the now famous cats whisker. Listening was done though a pair of headphones which meant that only one person could listen at a time.



Photograph Of a 1923 Radio

[from Pageant Of The Century - Odhams Press]

Later valve amplification could be added to radios which allowed them to power a loudspeaker so that everyone in a room could listen in.

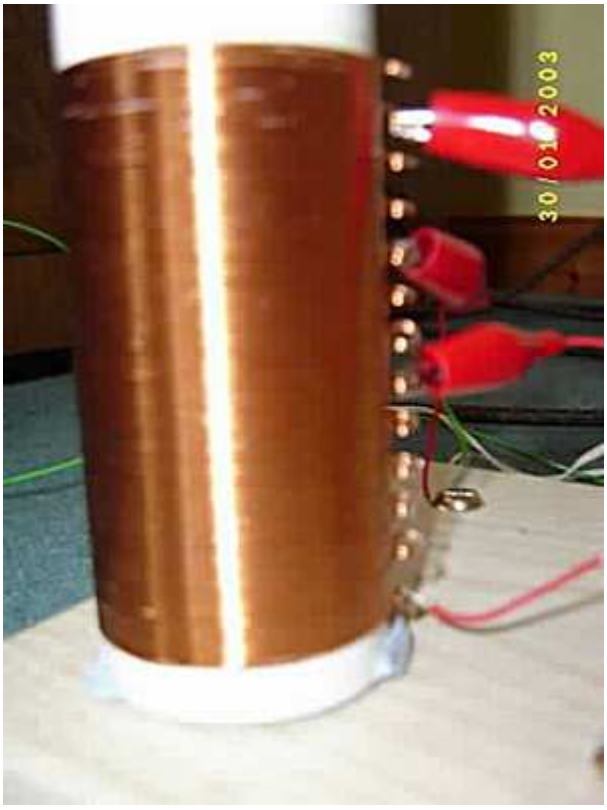
4 - HOW I BUILT MY CRYSTAL SET

The crystal set is a fun and easy thing to build.

The first thing I had to do was put up an aerial and earth. For the aerial I drilled a hole in the ceiling and fed wire through the hole. Then I went into the attic and took the wire to the other end of the attic, fastening it to rafters as I went. When I got to the other side I fed the wire out of an air vent to the outside of the house. I then went outside tied wire to a nearby tree and joined it up to the wire coming out of the air vent. This gave me an aerial of about 25 metres.

Next the earth. For this I drilled another hole in the ceiling and fed another wire through and out of an air vent at the other side of the house (see previous paragraph). I attached this wire to the metal frame of my swing. This is a good earth because its poles stick deep into the ground.

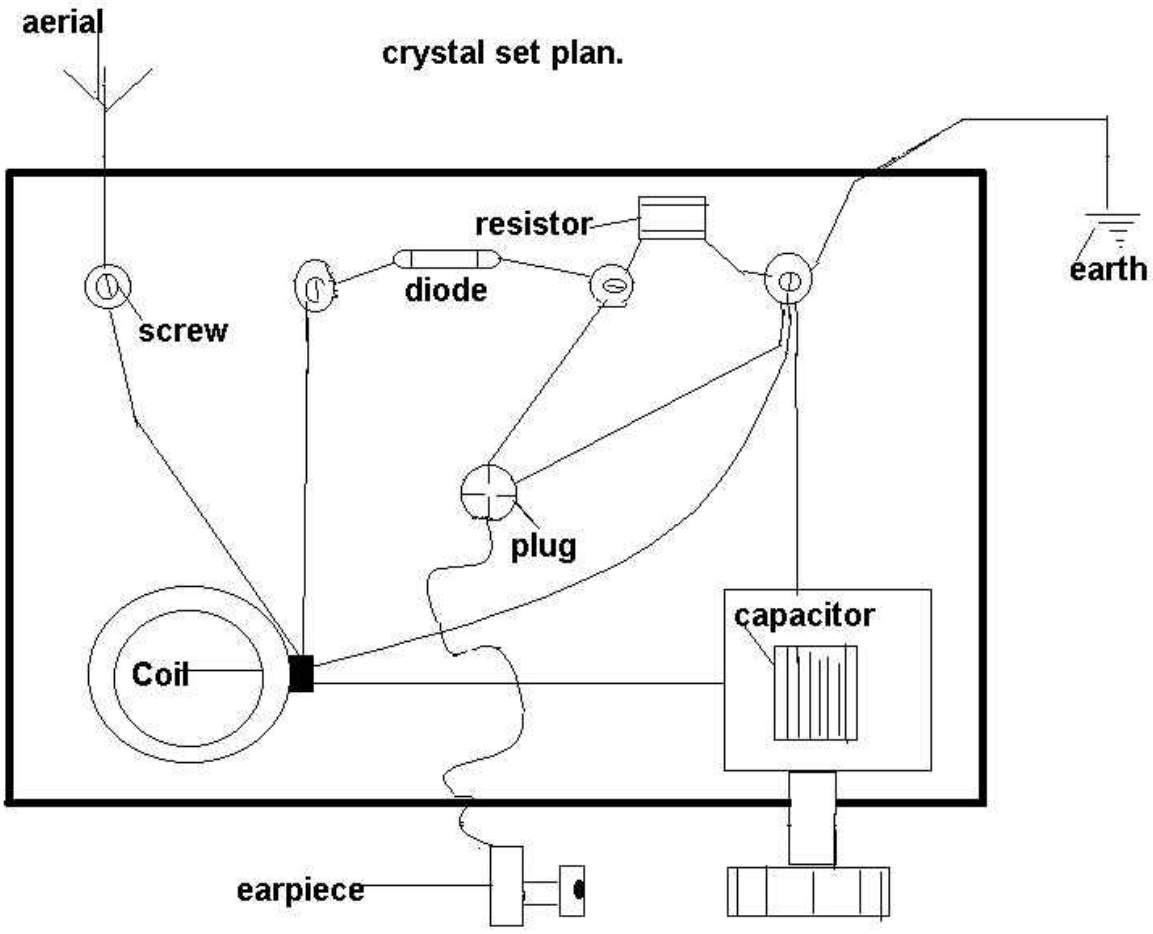
The coil used enamelled copper wire wound on to a piece of plastic pipe with a diameter of 5 cm. There were 120 turns of wire. Every 10 turns I made a loop on which a crocodile clip could be fixed. See the picture below:



Photograph Of The Coil Showing The Tapping Points & Crocodile Clips

The Tuning Capacitor - Being Carefully Adjusted

For the crystal set itself, I first had to plan what I would be using and where the pieces would be placed (see picture and the plan drawing). I was using the circuit shown above. I built the crystal set on a wooden board about 20cm by 15cm. I drilled 4 holes 4cm apart. I put screws with cup washers in each hole and fastened the wires of the parts of the crystal set and the aerial and earth wires under the cup washers. I fastened the capacitor to the board using Meccano brackets and screws. The coil was fixed to the board using glue. The earpiece plugged into a socket which was fastened to the board using a thunderbolt shaped piece of Meccano.

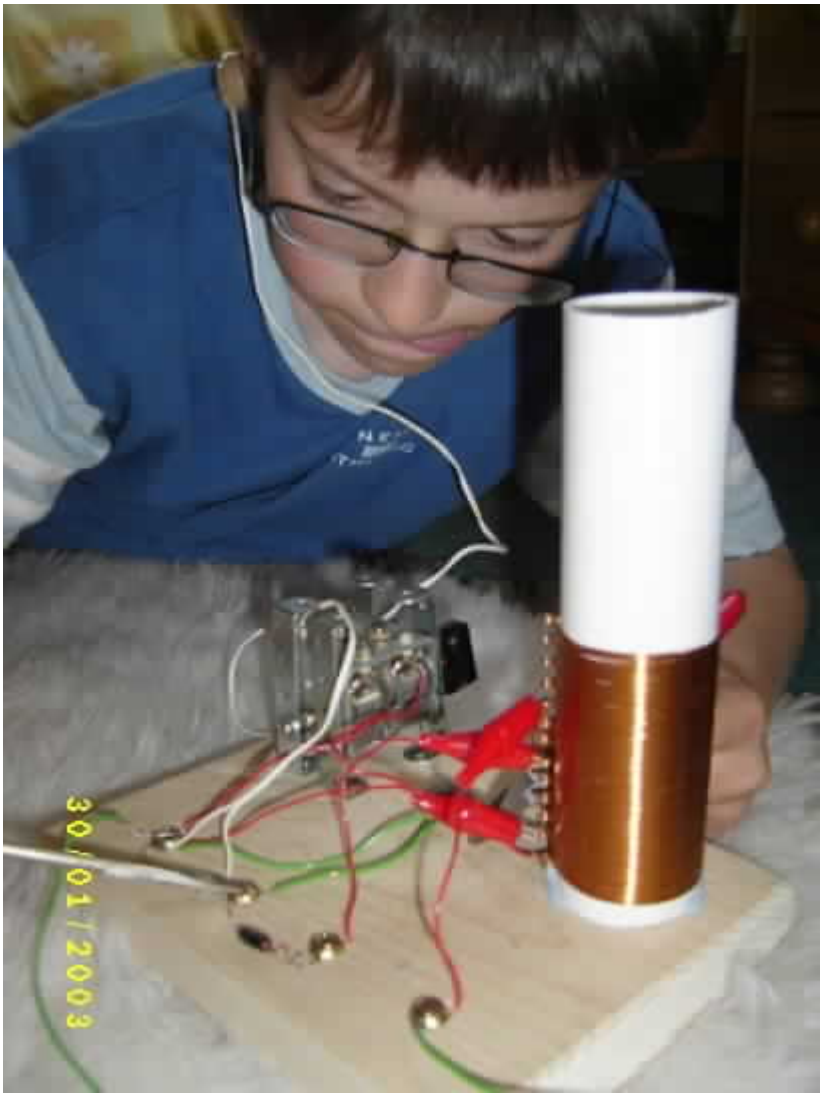


Drawing Showing Layout Of Components

5 - THE RESULTS I OBTAINED

At first all I could hear was a foreign station. But then, I tried late at night and could sometimes hear a station from Florida in English! By tuning the capacitor and moving the crocodile clips up and down the coil I managed to receive a few foreign stations on Short Wave. What I really wanted was to tune in to Medium Wave. I wanted to receive English stations like Radio Scotland and music stations. I tried everything including a longer aerial and using different parts but to no avail. I came to the conclusion that I could not get a strong enough Medium Wave signal in Northerly area of Scotland where we live.

Here is a picture of me operating my crystal set:



Photograph Showing The Crystal Set In Operation

6 - WHERE I FOUND THE INFORMATION

The History Of Crystal Sets

Book: The Pageant of the Century, Odhams Press, 1933.

www.oldradioworld.de/gollum/dhistor.htm

How I Built My Crystal Set:

Web page describing how to build a crystal set:

http://www.mds975.co.uk/Content/crystalsets2.html (THIS SITE)

Book:

"Making A Transistor Radio" written by Rev. G.C. Dobbs, Ladybird, 1972

Kits Are Available From:

www.midnightscience.com/kits.html

www.oldradioworld.de/gollum/xkits.htm

1N34A.com - crystal radio parts, including litz wire is http://1n34a.com

Other Hobby Websites:

Dave Schmarder's Crystal Set Pages http://www.makearadio.com/crystal/index.php http://www.makearadio.com

"Heart of England Crystal Radio Club" based in Birmingham, West Midlands - http://crystalradioclub.co.uk

More Crystal Set Links here >>

7 - QUIZ QUESTIONS

CRYSTAL SET QUIZ QUESTIONS:

1. What is a crystal set?
2. When were the first crystal sets sold?
3. Name 3 essential parts for a crystal set.
4. What does the earth do?
5. Does the coil pass signals of low frequency or high frequency?
6. How far apart were the screws on my crystal set?
7. What happened when I made the aerial longer?
8. What is this an anagram of: i a c c p r o t a?

(ANSWERS ARE FURTHER DOWN THIS PAGE)

PARTS THAT ARE REQUIRED

1	Wooden Base Panel
1	Coil Former 5cm dia. UPVC Pipe
1	Tuning Capacitor approx 500pF Air-Spaced or Solid Dielectric
1	Reel Of Enamelled Copper Wire for coil
1	Germanium Diode eg OA47 and IN34 which are especially suitable or other alternatives such as OA81, OA90 OA91, 1N94 etc
1	Crystal Earphone
1	47 k Ohm Resistor across earphone
6	½ inch No. 6 Brass Screws and Screw Cups
1	3.5mm mono Jack Socket
	Mounting Hardware (Meccano) for coil, capacitor and jack socket
	PVC covered wire for aerial and earth

A Superb RGD Valve Radio from 1937 - carefully restored in 1994 by Roddie Rankin

With thanks to Kenneth and Roddie Rankin for sending in this excellent article and photographs!

Crystal Sets Part 5 >

QUIZ ANSWERS

1. The first kind of radio
2. 1922.
3. Coil, diode, capacitor, earpiece, aerial, earth.
4. Gets rid of signals you don't want.
5. Low.
6. 4 cm.
7. It did not make much difference.
8. Capacitor.

No AM radio stations or transmitters in your locality or country?

Has your local medium wave broadcast station closed or been moved to VHF/FM or Digital? Don't worry. You can still build and experiment with crystal sets and TRF radios by also buying or even building a simple low power AM transmitter. So, not only can you use your crystal sets but you can also run your own radio station that can be heard in and around your home - playing the music or programmes that you want to hear!

SSTRAN AMT3000 : Superb high fidelity medium wave AM transmitter kits from SSTRAN. Versions available for 10kHz spacing in the Americas (AMT3000 or AMT3000-SM) and 9kHz spacing in Europe and other areas (AMT3000-9 and AMT3000-9SM). Superb audio quality and a great and well designed little kit to build: http://www.sstran.com/pages/products.html

http://www.sstran.com/

Other AM transmitters available:

Spitfire & Metzo: Complete, high quality ready built medium wave AM Transmitters from Vintage Components:
http://www.vcomp.co.uk/index.htm Vintage Components offer a choice of the high quality Spitfire and Metzo transmitters:

SPITFIRE AM Medium Wave Transmitter with 100 milliwatt RF output power:

http://www.vcomp.co.uk/spitfire/spitfire.htm

METZO AM Medium Wave Transmitter with built in compressor:

http://www.vcomp.co.uk/metzo/metzo.htm

AM88 LP A basic AM transmitter kit from North County Radio.
http://www.northcountryradio.com/Kitpages/am88.htm

LINKS:

BOWOOD ELECTRONICS - A friendly, helpful and very speedy source for many of your electronic components at prices that won't frighten your wallet!

THE FOXHOLE and P.O.W RADIOS - Simple crystal set receivers used by soldiers during the war and by prisoners of war (P.O.W.'s).

VINTAGE COMPONENTS - A great resource for crystal sets, components, valve radio kits and medium wave AM transmitters!

6V6 - Electronic Nostalgia and Vintage Components

MORE CRYSTAL RADIO LINKS HERE >>

Crystal Sets (Part1) | Build Your Own Crystal Set (Part 2)

Spider's Web Crystal Set (Part 3) | Experimental Crystal Sets (Part 5) | Crystal Radio Links

^Top Of Page

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What Do the Markings on Circuit Breakers Mean?

by Ed Larsen, P. E.

Have you ever been confused about what the markings on circuit breakers mean? Understanding the markings on electrical equipment is a fundamental need to ensure a safe and reliable electrical installation. Circuit breaker marking requirements are established by the requirements found in the *NEC* and the UL 489 product standard. This article will discuss the most common markings and where they can be found.

The UL 489 product standard for Molded Case Circuit Breakers specifies the information to be marked on circuit breakers and where it is to be located, so let's discuss what information needs to be marked on the circuit breaker and the location where you will find those markings. Keep in mind the UL® standard specifies minimum requirements. Circuit breaker manufacturers may provide additional information or provide information in a more convenient location.

Markings Visible without Removing Trims or Covers

UL 489 requires that some markings be visible without removing trims or covers. This location is typically referred to as the handle escutcheon (see photo 1).

Markings Visible with Trims or Covers Removed

UL 489 requires other markings be visible on an installed circuit breaker with trims or covers removed. This location is typically referred to as the face of the circuit breaker (see photos 2–4).

Other markings which should be visible with trims or covers removed are:

Independent trip – Multi-pole circuit breakers are constructed with either a common trip, where all poles are mechanically tripped when one of the poles trips, or an independent trip construction where only the pole that is involved with the overcurrent condition trips. If a 2-pole circuit breaker does not have an internal com-

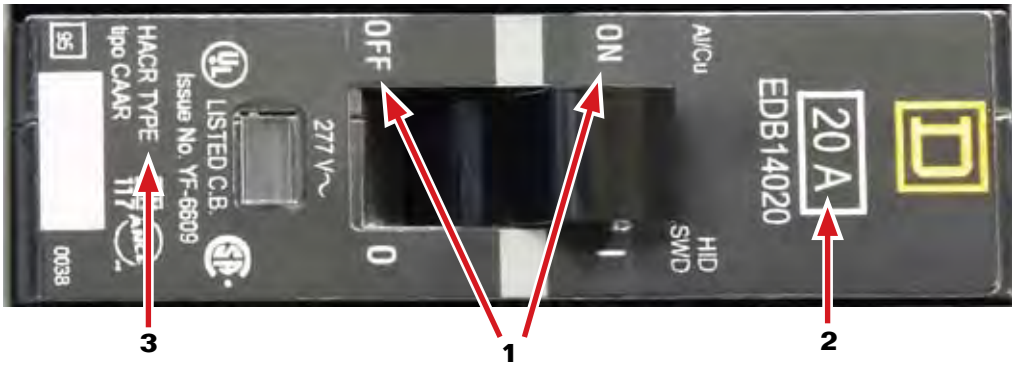

Photo 1. Markings Visible with Trims or Covers in Place

1. ON and OFF – The ON and OFF (closed and open) positions of the handle must be marked (*NEC* 240.81). These positions may also be marked with the internationally recognized "I" and "O" symbols, although this is not a UL requirement. If these markings are not visible when a motor operator is installed over the circuit breaker markings, then they must appear on the motor operator. Motor operators may be found in applications where remote or automatic operation of a circuit breaker is required.

2. Ampere rating (if 100 A or less) – The ampere rating may be located on the handle escutcheon or on the handle itself (*NEC* 240.83(B)). Circuit breakers that are rated more than 100 A may have their ampere rating marked in a position that is not visible with trims or covers in place.

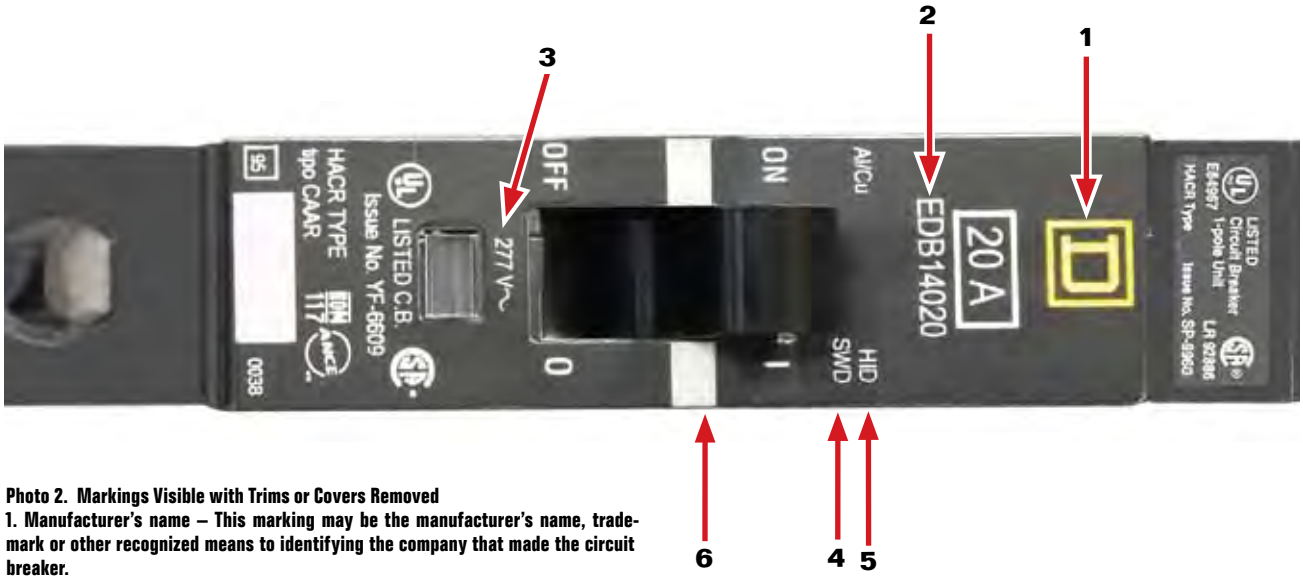

Photo 2. Markings Visible with Trims or Covers Removed

1. Manufacturer's name – This marking may be the manufacturer's name, trademark or other recognized means to identifying the company that made the circuit breaker.

2. Type designation – All circuit breakers are marked with a type designation, which may be a catalog number prefix or a separate designation. Equipment labels, such as on panelboards, will list the circuit breaker types suitable for use. Note that the word "type" may or may not be used on the circuit breaker or equipment labels. It is important to review the markings on the equipment, such as a panelboard, to make sure the circuit breaker designations on the equipment match the marking on the circuit breaker.

3. Voltage rating – All circuit breakers must be marked with a voltage rating. If the rating is not marked "ac" or "dc," then it is suitable for both. 120/240 V rated circuit breakers are suitable for use on single and three-phase 4-wire systems where the line-to-ground voltage does not exceed 120 V. Wye rated circuit breakers such as those rated 480Y/277 V, are suitable for use on three-phase 4-wire systems where the voltage to ground does not exceed 277 V. Special attention needs to be given to high leg or corner-grounded delta systems to insure that the circuit breaker has the appropriate rating. A review to see that the circuit breakers installed have a voltage rating suitable for the application is paramount for a code-compliant installation (*NEC* 240.83(E)).

4. SWD – 15- or 20-A circuit breakers rated 347 V or less may be marked "SWD," meaning that they are suitable for switching fluorescent lighting loads on a regular basis (*NEC* 240.83(D)). These circuit breakers are evaluated for high endurance use, since they will be used similar to a light switch.

5. HID – 50 A or less circuit breakers rated 480 V or less may be marked "HID," meaning they are suitable for switching high intensity discharge or fluorescent lighting loads on a regular basis. These circuit breakers may employ a different construction than a standard SWD circuit breaker in order to address the high inrush current resulting from the lower power factor created by the HID lighting (*NEC* 240.83(D)). These circuit breakers also undergo additional endurance evaluation to demonstrate their ability to perform the switching duty.

6. Trip and reset – Circuit breaker handles typically assume an intermediate position when tripped. This position must either be marked on the circuit breaker or on the equipment into which it is to be installed. If these markings are not visible when a motor operator is installed, then a "tripped" marking may appear on the motor operator.

mon trip feature, then it must be marked "Independent Trip" or "No Common Trip." *NEC* 240.20(B) is the foundational requirement for a common trip function in a circuit breaker; however, it also goes on to explain where independent trip is permitted.

For Replacement Use Only not-CTL – The Class CTL (circuit limiting) panelboard has only been in existence for about 25 years, even though the lighting and appliance branch circuit panelboard has been in the *NEC* for decades. CTL panelboards have a rejection means designed to reject more than the appropriate number of circuit breakers that can be installed in the panel. The marking "For replacement Use Only Not CTL Assemblies" means that the circuit breaker does not have CTL rejection provisions and is intended for replacement in older equipment pre-dating the CTL requirements for circuit breakers and panelboards. Circuit breakers with this designation should not be installed in a panelboard marked "Class CTL Panelboard" since that would be a violation of the listing of the assembly [*NEC* 110.3(B)].

Markings Found in Other Locations

The markings we will discuss below may appear in any location except the back of the circuit breaker. These markings include:

40°C – This marking indicates the maximum ambient temperature in which the circuit breaker can be applied at its marked ampere rating without rerating the ampacity of the circuit breaker. This marking is required for thermal-magnetic circuit breakers and is optional for electronic trip circuit breakers unless they are only suitable for a 25°C ambient, in which case they must be marked 25°C. When the ambient temperature rises above 40°C, the designer may need to consult the manufacturer to obtain rerating information (see item 4 in photo 3).

Class CTL – Circuit breakers marked Class CTL have a rejection means designed into the circuit breaker. Class CTL panelboards or assemblies, in conjunction with Class CTL circuit breakers, prevent more circuit breaker poles from being installed than the number for which the equipment is rated.

Photo 3. Markings Visible with Trims or Covers Removed

1. Line and load designation – Circuit breakers marked with "line" and "load" designations are not suitable for reverse connection. Circuit breakers with interchangeable trip units must be marked "line" and "load" unless there is no risk of shock when changing the trip unit.

2. Interrupting ratings – All circuit breakers with an interrupting rating more than 5000 A must be marked with an interrupting rating (*NEC* 240.83(C)). Interrupting ratings are stated in RMS symmetrical amperes. If the short-circuit current rating of the equipment in which the circuit breaker is installed is less than the interrupting rating of the circuit breaker, then the lesser rating applies. Circuit breakers should be reviewed after installation to ensure they have an interrupting rating suitable for the application. This marking may be found in any location except the back of circuit breakers that are 1-½ inches wide per pole or less due to the size constraint.

3. Ampere rating (if more than 100 A) – The ampere rating of a circuit breaker larger than 100 A may be found in a location that is visible after the cover or trim is removed. This marking requirement also applies to interchangeable trip units (*NEC* 240.83(A)).

5. Terminations (Cu-Al wire) – Circuit breakers must be marked with the type material (Cu-Al) and size of wire for which their terminals are suitable for use. If suitable for use with only copper or only aluminum, then the word "only" must be used. The abbreviations "CU" and "AL" are generally found on circuit breakers as permitted by the product standard. If only solid 10–14 AWG wire can be used, then that information must be noted. The number of wires per terminal will also be noted if more than one wire per terminal is permitted (*NEC* 110.14(A)).

6. Tightening torque – The nominal torque for all field-wiring terminals must be marked. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be found in any location except on the back.

7. Wire temperature ratings – Circuit breakers rated 125 A or less may be marked as suitable for use with 60°C, 60/75°C or 75°C only wire. Circuit breakers rated more than 125 A are rated for use with 75°C wire; the marking is optional. It is always permissible to use wire with a higher temperature rating, but it must be sized in accordance with the temperature marking on the circuit breaker and *NEC* Table 310.16. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be in any location except on the back (*NEC* 110.14(C)1).

8. HACR type – This marking indicates the circuit breaker is suitable for use with the group motor installations typically found in heating, air conditioning and refrigeration equipment. The *NEC* 2005 no longer has this marking requirement. The electrical industry determined that circuit breakers are considered suitable for use with such equipment without any further testing, therefore, the HACR marking is no longer required on air conditioning and refrigeration equipment or on circuit breakers for use in these applications. The requirement for this marking has also been removed from the UL 1995 product standard for HVAC equipment (see item 3 in photo 1).

Maximum wire size – Circuit breakers are typically marked with a wire range, however that marking is not mandatory. If the circuit breaker cannot accept the next larger wire size required for the ampere rating, then the maximum wire size must be marked in any location except the back (see item 5 in photo 3).

Separately shipped connectors –If connectors are not factory installed on a circuit breaker, then it must be marked with the proper connectors or terminal kits required in any location except the back (see item 8 in photo 3).

Ground-Fault Protection for People

The GFCI function, as part of a circuit breaker, provides ground-fault protection for people and has a number of unique marking and instruction requirements.

Test function – The GFCI has a test function that requires action upon installation and on a monthly basis. GFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate monthly testing.

Photo 4. 1Ø – 3Ø Marking
1. 1Ø – 3Ø marking – A 2-pole circuit breaker used to protect a 3-phase load on a corner-grounded delta system must be rated and marked for such an installation. Circuit breakers marked "1-phase – 3-phase" or "1Ø – 3Ø "are suitable for use on 3-phase corner-grounded delta or single-phase circuits (*NEC* 240.85).

Suitable for continuous operation at 100% of ampere rating only if used in a recognized enclosure 52 in. (1321 mm) by 20 in. (508 mm) x 7 1/2 in. (190 mm) deep or larger. Use only with 90° rated wire sized per the ampacity of 75° rated wire.

Figure 1. 100% Rated Marking
100 percent continuous rated – Circuit breakers are typically intended for use at not more than 80% of rated current where the load is considered continuous, or will continue for 3 hours or more (*NEC* 210.20). However, some circuit breakers are rated for continuous use at 100% of their current rating. These circuit breakers must be so marked in any location except on the back. Enclosure information such as a specific type or specific volume must also be marked. A requirement for the use of 90°C insulated wire sized to the 75°C column in *NEC* Table 310.16 and specific ventilation requirements may also be marked on the circuit breaker or equipment (*NEC* 210.20(A) and 215.3).

"Class A" marking – A "Class A" ground-fault device is intended to protect people. The Class A marking indicates that the trip threshold of the GFCI is between 4 mA and 6 mA. This marking may be in any location except the back.

Instructions – All GFCI circuit breakers must include instructions for the installer plus instructions on the use of the test function. A hangtag or self-adhesive label must also be provided, instructing the user to test the GFCI at least monthly. Inspectors should check to see that the tag or label has been properly installed.

Ground-Fault Protection for Equipment

Circuit breakers may also include a ground-fault protection for equipment (GFPE) function that, like GFCIs, has a number of unique marking and instruction requirements.

Test function – GFPE circuit breakers may have a test button or switch that may be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Trip level – GFPE circuit breakers must be marked with their trip threshold in milliamperes in a location accessible without removing trims or covers.

Instructions – All GFPE circuit breakers must include instructions for the installer.

Arc-Fault Protection

Circuit breakers may also include arc-fault protection (AFCI) that, like GFCIs, also has a number of unique marking and instruction requirements.

Device identification – AFCIs must also be identified appropriately. Branch/feeder or Combination type AFCIs must be so marked in a location visible when the trims or covers are removed. This is an important marking to note as we move into 2008, as *NEC*-2005 requires Combination AFCIs in bedrooms effective January 1, 2008 (*NEC* 210.12).

Test function – AFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Instructions – All AFCI circuit breakers must include instructions for the installer.

Circuit Breaker Markings Ensure a Safe Electrical Installation

So why are all of these markings on circuit breakers? Without them, it would be nearly impossible to install or inspect an installation for the appropriate performance ratings and fundamental electrical connections. When designing or completing an installation, key items to review are:

1. Are the voltage, continuous current, and interrupting ratings appropriate for the application?
2. Does the application require SWD or HID ratings?
3. Is the wire type and size appropriate for the circuit breaker?
4. Is the circuit breaker suitable for the equipment in which it is installed? Have other protective functions such as GFCI or AFCI been provided as required by the *NEC*?
5. Is the temperature rating of the circuit breaker suitable for the application?

The UL Marking Guide for Molded Case Circuit Breakers is a valuable resource to understand circuit breaker markings that may further explain these and other markings in detail. If you have questions about CB markings not answered here, consult the Marking Guide or the manufacturer to assist in an *NEC*-compliant installation.

Mr. Larsen, P. E., is the manager, Industry Standards for Circuit Protection for Square D Company / Schneider Electric. Mr. Larsen has served in various positions over the past 32 years. He is a member of several UL, IEEE and NEMA technical committees. His other memberships include NFPA, IAEI and CANENA.

Magnetic Loop Antenna's.

Why is an Magnetic Loop antenna so special, this antenna is picking only the MAGNETIC part of the ELEKTRO MAGNETIC radio wave. The big advantage of this antenna is that the electric interference from the big city (streetlights, television's , cars etc...) have no influence on the received signal. With the loop you can hear other stations that you can't hear if you use a DIPOLE, with a dipole the stations are buried in the noise

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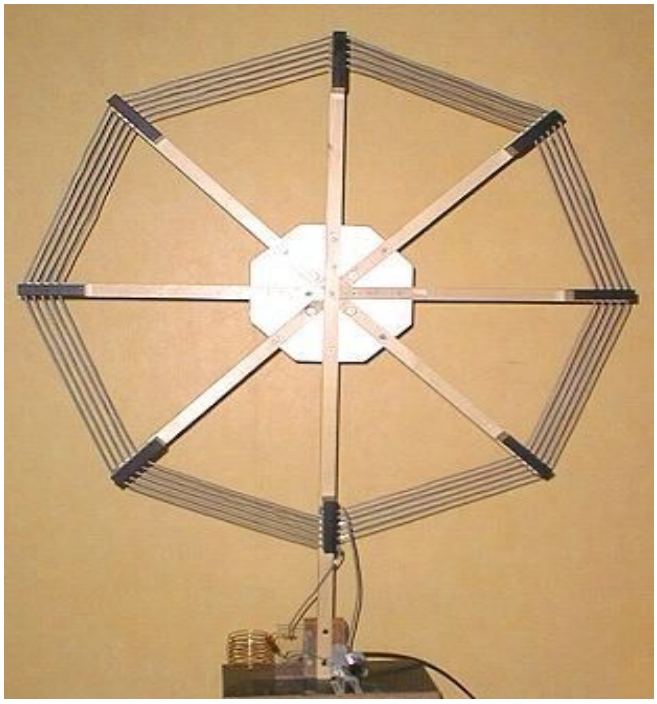
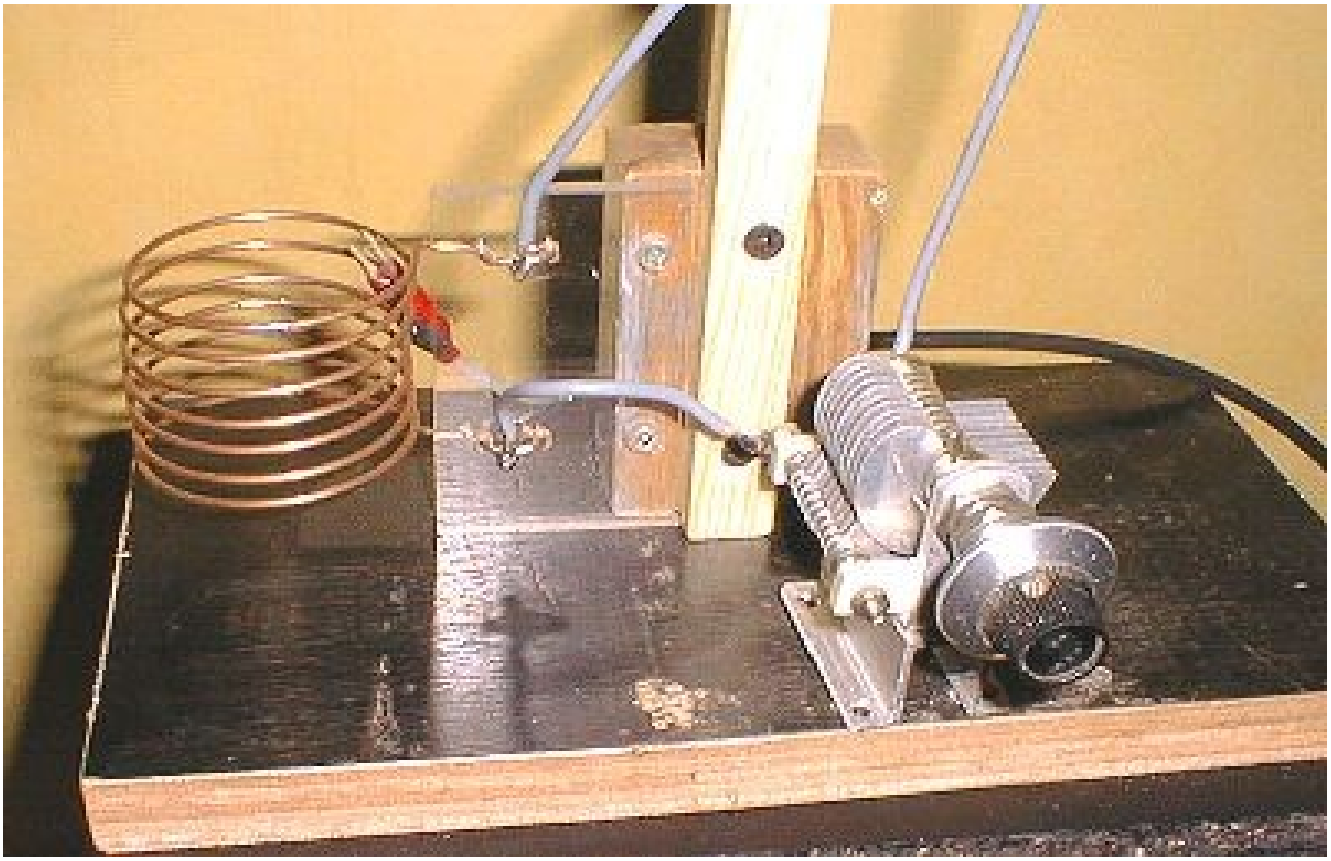
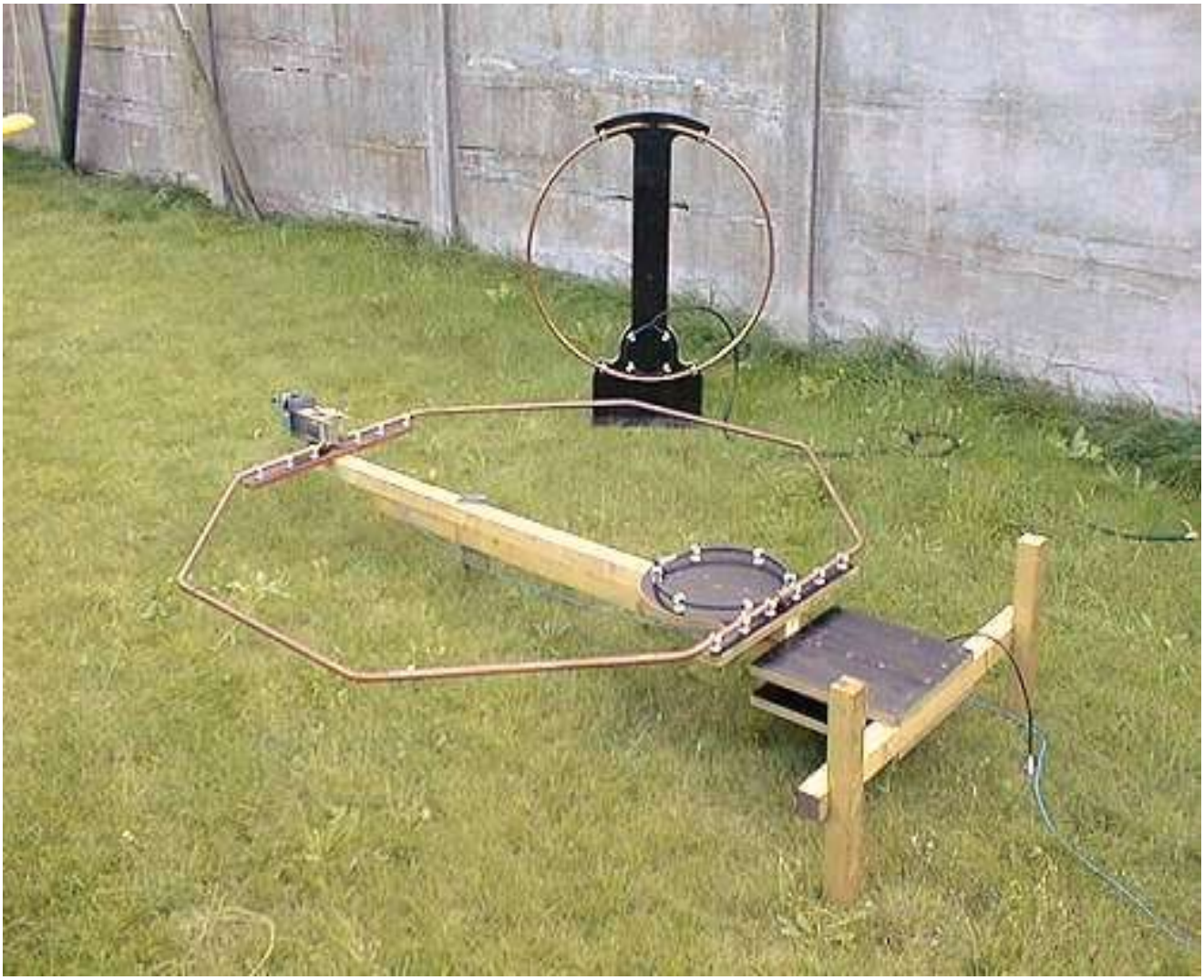
Mutil-turn Magnetic Loop

This is the first loop I build from a article in the QST from February 1996, it's 30 Inch-diameter, and it's designed by G2BZQ/WØ for 80 M



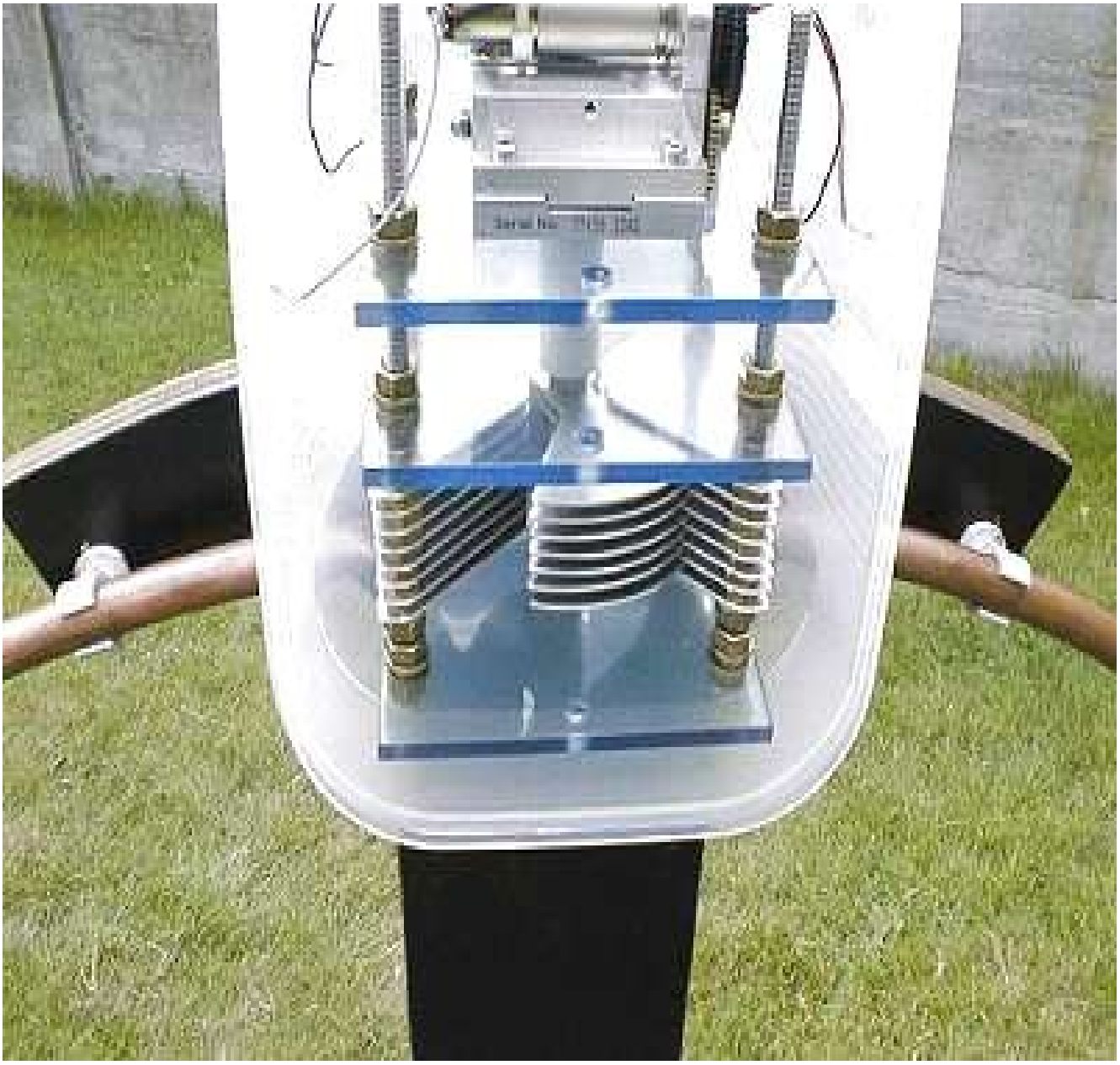


Single Turn Magnetic loops



The first one turn loop that I built was made from 75 Ohm TV Coax and with a small explanation in the RSGB handbook for radio amateurs. I used the outer screen from the coax and the results of the loop where good.The next loop I built is a octagon loop in 15mm copper tube with a circumference of 4.8 meter (16 feet).
The frequency range of this loop goes from 14 MHz to 7 MHz and works fine. The biggest problem is the tuning capacitor, if you transmit with a power of 100 W you need a capacitor with a voltage rating of 5000 Volt.

A capacitor which can handle this voltage is hard to find over here and if you find one they are very expensive. The first capacitor I built was a design from GW3JPT from a article in the RADIO COMMUNICATION from February 1994. It is a split stator capacitor with a capacitance of 140 pF and with a voltage rating of 6000 Volt. The capacitor is remote tuned with the use of a small BBQ spit motor



The second capacitor I built is my own design and it's a butterfly capacitor because the losses are lower than a split stator. The capacitance is 5-65 pF and the voltage rating is 7200 volts. I used it for the small loop with a dia. of 800 mm (2.66 feet) and the frequency range of this loop is from 28 MHz to 14 MHz. The Aluminum plates of 1 mm for the capacitors are cut with a JIG SAW

FAQ

I'd like to talk a little more on your setup. it seems like something which I could get together if only some more data was available. do you have any notes etc still laying about since its build ?

The theory for calculating the loop is very simple. The circumference of a magnetic loop is 1/4 wave of the design frequency.

Example for 14 MHz.

300 / 14 MHz = 21.428 m is 1 wave
21.428 / 4 = 5.357 m is 1/4 wave circumference
5.357 / 3.14 = 1.706 m diameter.

The recommendations are that you can tune the loop from the design frequency to the frequency divided by 2 to keep the efficiency acceptable.

14 MHz / 2 = 7 MHz

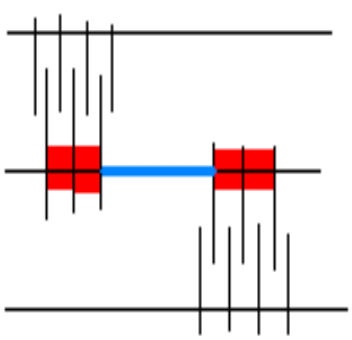
I made the small loop (800mm / 31.5 ") from soft copper tube on a role that you can buy in a plumber shop and it's easy to make a nice circle if you draw on the ground a circle with a rope and a piece of chalk.

For mounting the loop to the hardboard I used plastic clamps that they use for mounting copper tube on the wall.

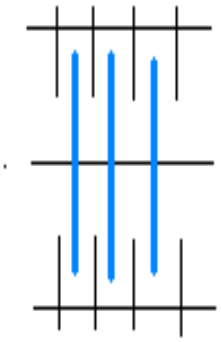
Why is a butterfly capacitor better?

For high voltages and currents the use of Capacitors with wiper contacts is not recommended. That's why they use capacitors in series. The pro for series's capacitors is that the voltage rating is doubled. The anti is that the value of capacitance is divided by 2. For the split stator capacitor the 2 capacitors are connected in series by the shaft (blue) and the red spots on the first drawing are losses.

For the butterfly capacitor the 2 capacitors are directly connected in series by the rotors and gives less losses.



Split stator



Butterfly

Do you know of anyone that has built a similar loop that outperformed a garden variety dipole?

Comparing antennas is very difficult , sometimes I have heard 60 % better signals in RX and TX on the loops then on the dipole.

In theorie the performance of a magnetic loop is - 0.4 dB lower then a dipole or a vertical .

I have a homebrew trap dipole from 40-20-15-10m and the height above ground is only 7 m(23 ft). For a good performance on 40 m the dipole must have a height of 1/2 wave above ground (66ft). I don't have a radiation angle on 40 m and it's only good for contacts in Europe and not good for DX, now the 1.5 m loop tuned to 7 MHz with a efficiency of 38 % (38 w ERP) and a angle radiation of about 20 degrees performs better than the dipole because the vertical magnetic loop only 1 M above ground as a angle radiation and the dipole don't.

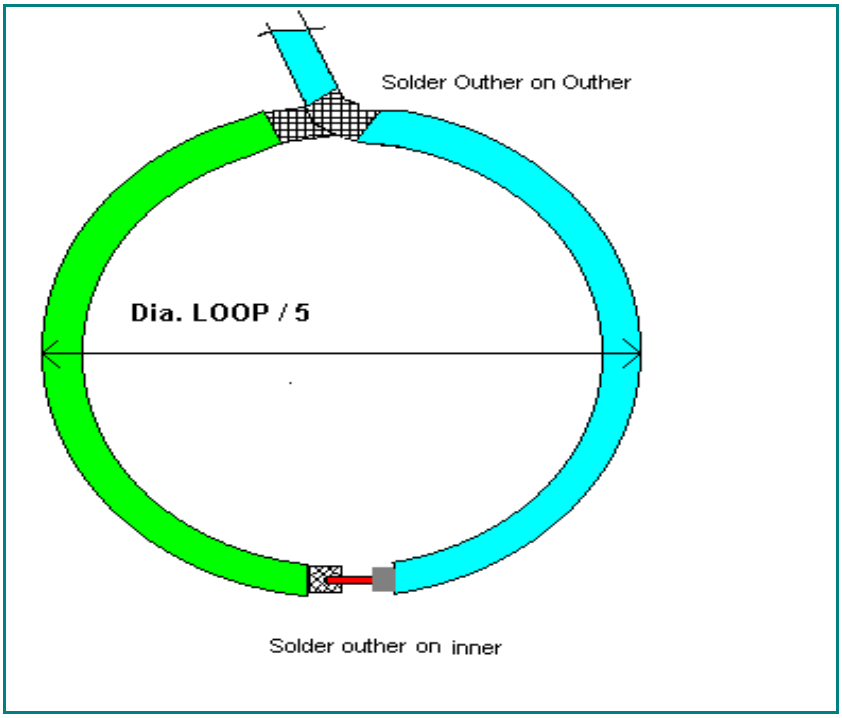
Another advantage is that the reception on a loop is much better, on 20m I have with the dipole S5 noise from the big city, if I switch to the loop I have S1 noise and hear stations who are buried in the noise when I use a dipole.

Coupling loop dimensions?

I find that the best way to feed the loop is with the shielded 1/5 Faraday loop made from coax RG213 or RG8, I tried the gamma match bud I had problems to keep the VSWR low on all Bands, the shielded loop gives on all bands VSWR 1.1 and reduce more noise pick-up then the gamma match.

I found out that if you use a 1/5 Faraday loop, that the loop is to big, making the loop smaller with 0.5 inch by the time in circumference and checking with a field strength meter you can see that the radiated power increase.

The place off the feeding loop is placed at the electrically neutral point on the loop and that is 180° from the capacitor and I have the best results with the feeding loop close to the ground and the capacitor far from the ground.



I was wondering if you worried about the resistance of the mechanical joints (copper pipe bolted to the capacitor) significantly reducing your radiation efficiency as I think the radiation of these antennas can get as low as .01 ohms?

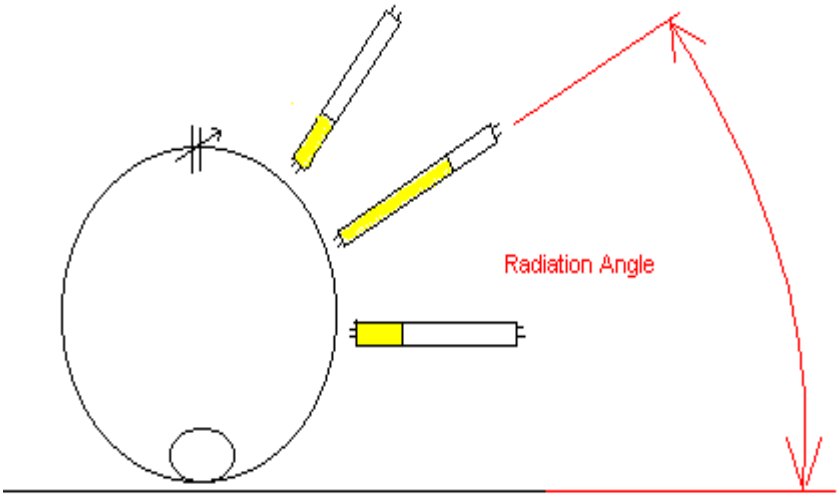
Solder or weld the capacitor plates is always the best, but I'm afraid if you make the spacers and the plates in ALU that with welding everything is going to bend from the heat and I know from practice (I work in a maintenance workshop) that welding ALU is coarse. Another possibility is using all brass or copper and solder, there are hams that using double PC board for the plates.I made a QSO in phone with Florida, RPRT 5-5 and the other station used a vertical antenna, with the small loop (800mm and theoretical efficiency 41 % on 14 Mc) vertical in the garden and the states side is through the house. I was very happy with the results , so I think that a capacitor made with torqued compressed joints is good enough for using 100 W.

Have the dissimilar metal joints weather well?

To keep the oxidation low on the dissimilar metals I used a thin coat of Vaseline after assemble the capacitor and with the Tupperware a like plastic box it is good protect against all weather conditions.

How to find the radiation angle of the antenna? Can it be found practically?

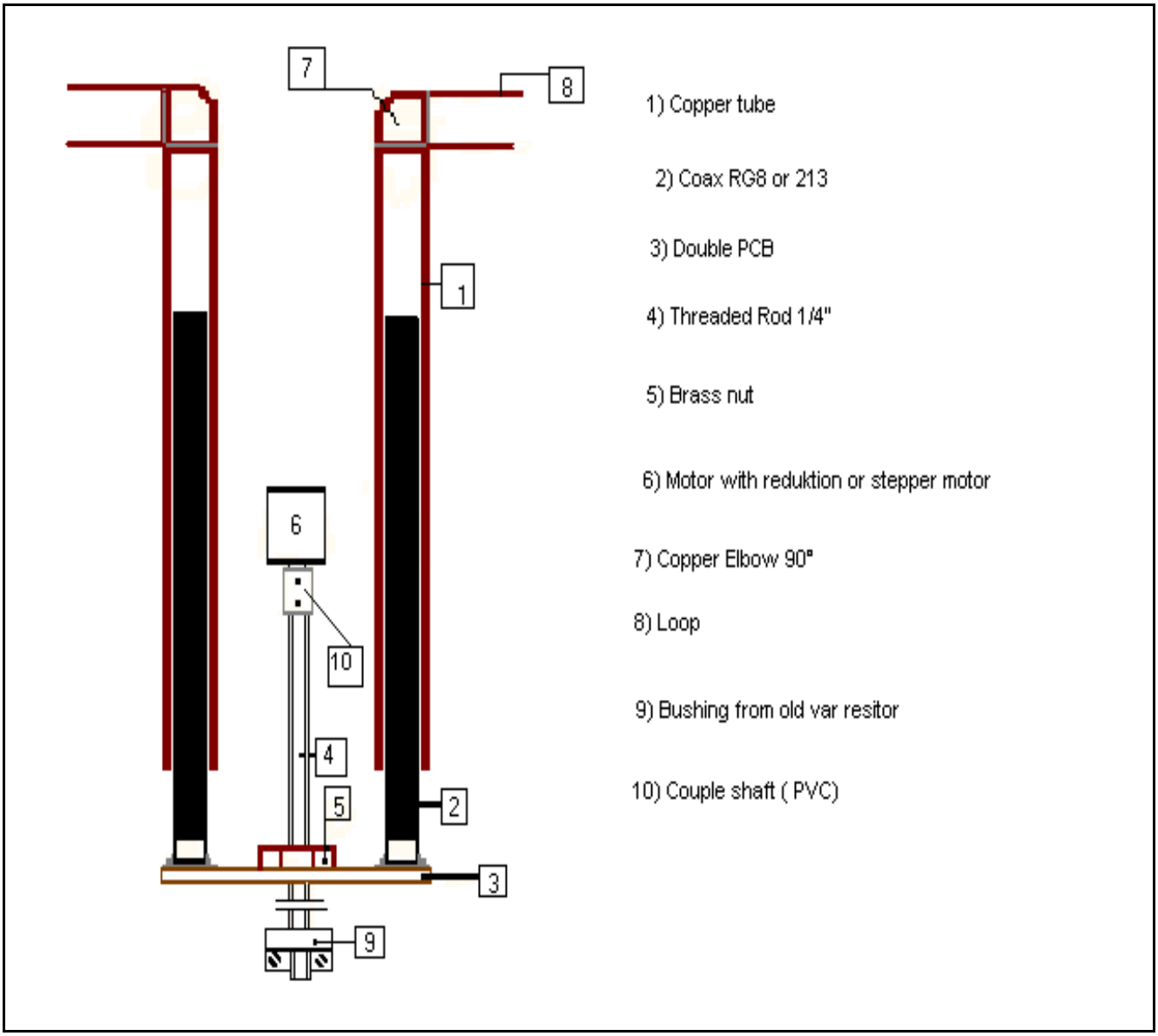
Finding the radiation of a magnetic loop is very easy, with a TL-light tube you can see it, with about 10 w power on the the loop with the TL-tube in the plane of the loop at right angle to the circle you see the tube lighting, there where the the light is the farthest on the tube thats the radiation angle.



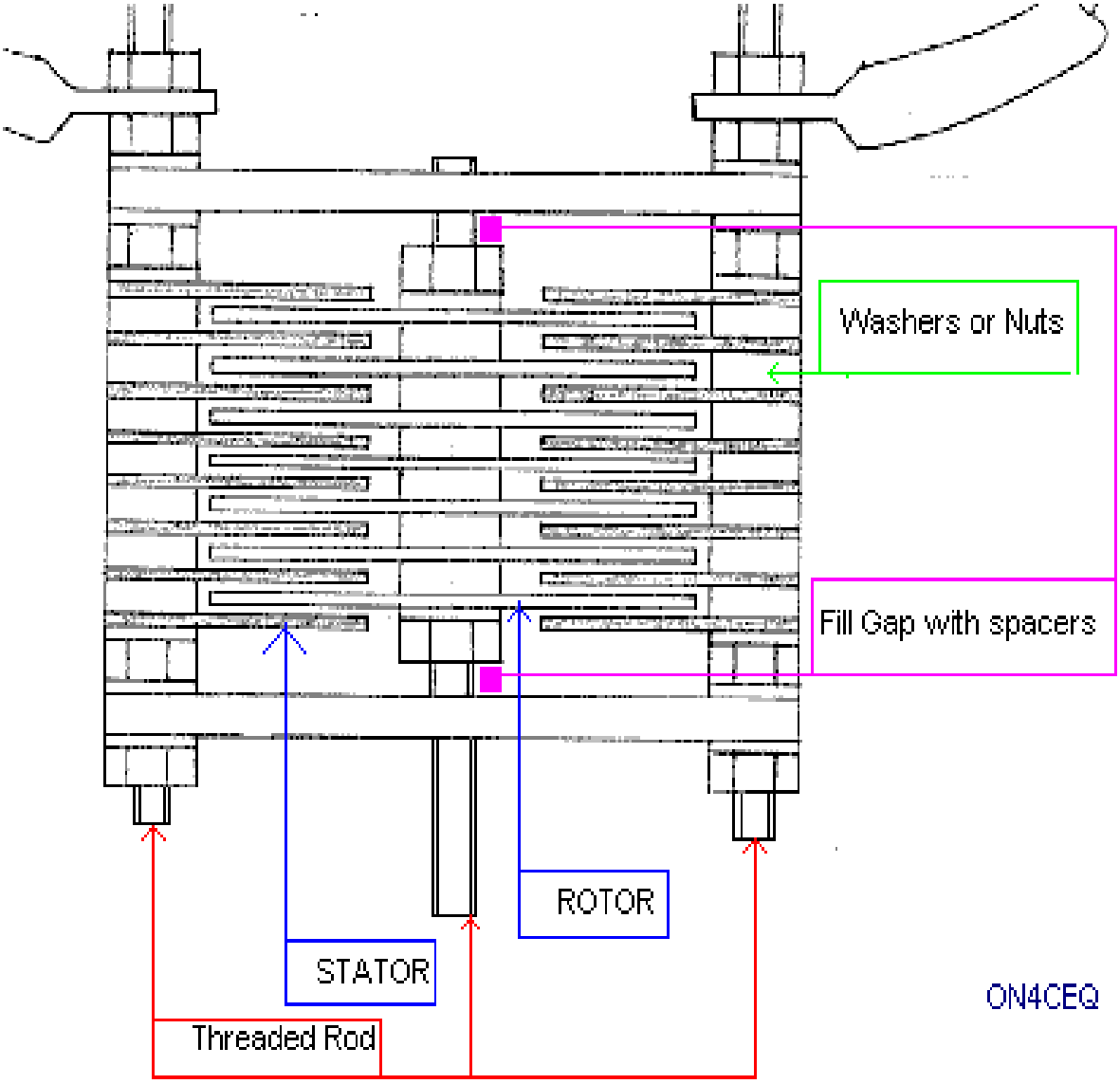
When you refer to washers, nuts and rods you use the term "M6".Please forgive my ignorance, but to what does "M6" refer? Does this mean 6mm?

M6 is (M=metrical) and 6 is indeed 6 mm threaded rod and you can compare the size with W1/4" (6.35mm)

A very easy to build Piston Capacitor.



How to build your own Butterfly Capacitor.



The best material for the front and the back is CLEAR PVC 3 or 5 mm thick as alternative you can use GREY PVC or 2 sheets pc board together with the copper removed .

The best material for the washers, nuts (M6) and threaded rod (M6) is brass or stainless steel,(NON MAGNETIC MATERIALS for the losses).

For the spacing of the vanes you can use 2 washers M6= (6Kv) or a nut M6 =(12 Kv) if you use aluminum plate 1 mm thick.

If you use a nut then the best thing to do is remove the thread by drilling withØ 6.2 mm.
The effective area for the vanes is 11.7 cm² and with the formula

for 2 washers = (0.0885 x 11.7 cm²)/ 0.1 cm = 10.35 pF for 1 air gap.

for 1 nut = (0.0885 x 11.7 cm²)/0.2 cm = 5.17 pF for 1 air gap

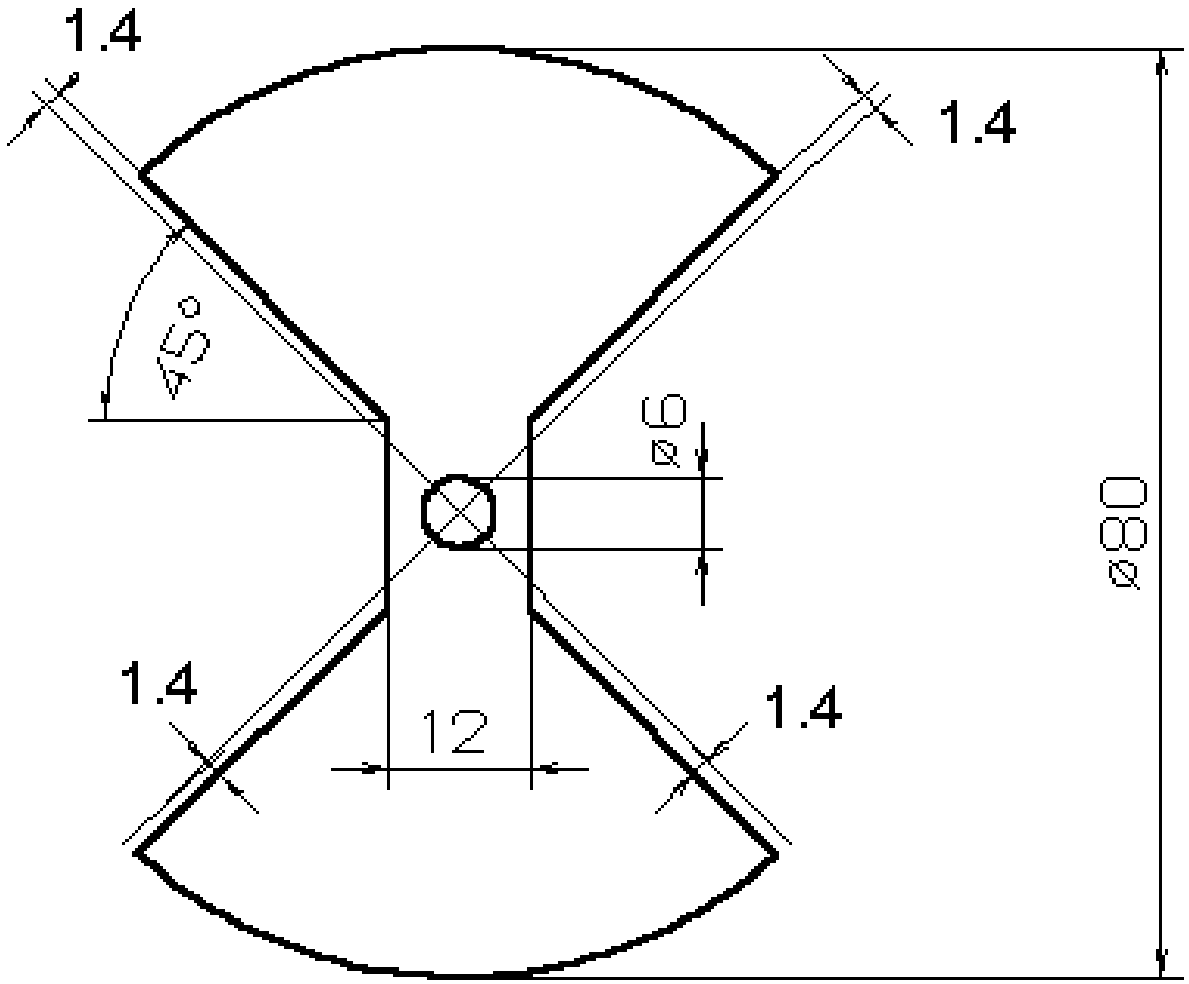
Example:

If you you make a capacitor with 2 washers as spacing and you make
5 rotor vanes and 6 stator vanes then you have 10 air gaps.

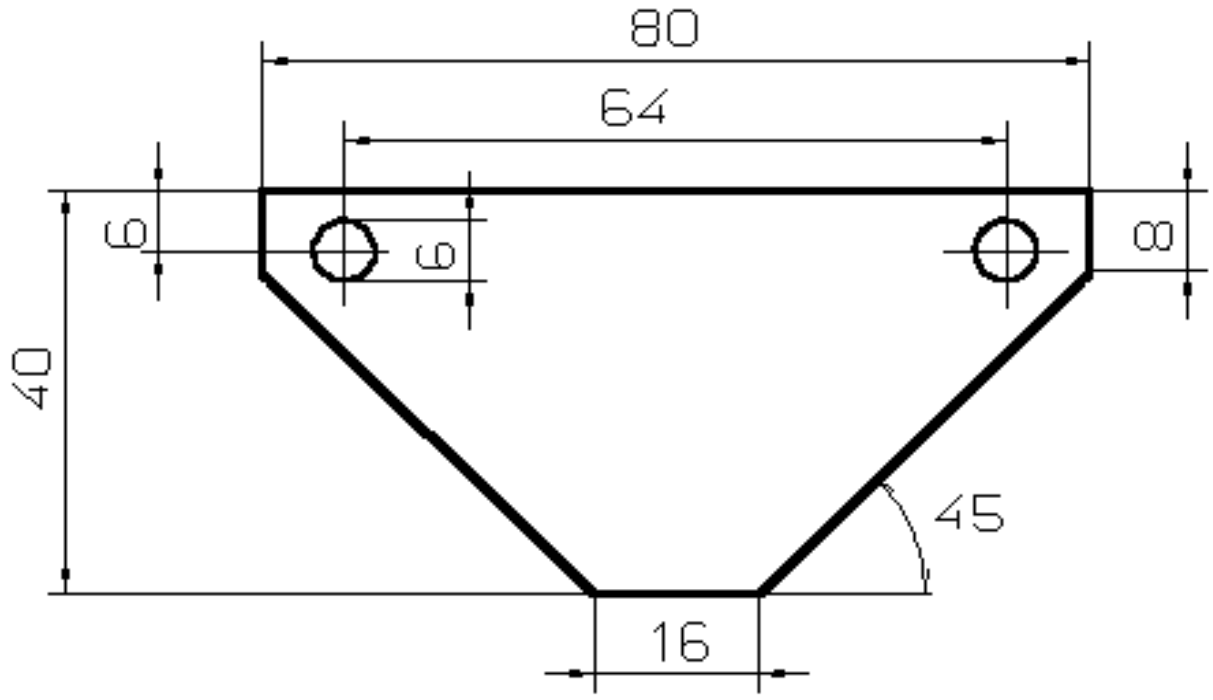
10.35 pF x 10 = 103 pF + 10 pF stray capacitance = 113 pF / 2 = 56 pF

The final result is a capacitor with a value from 5 - 56 pF.

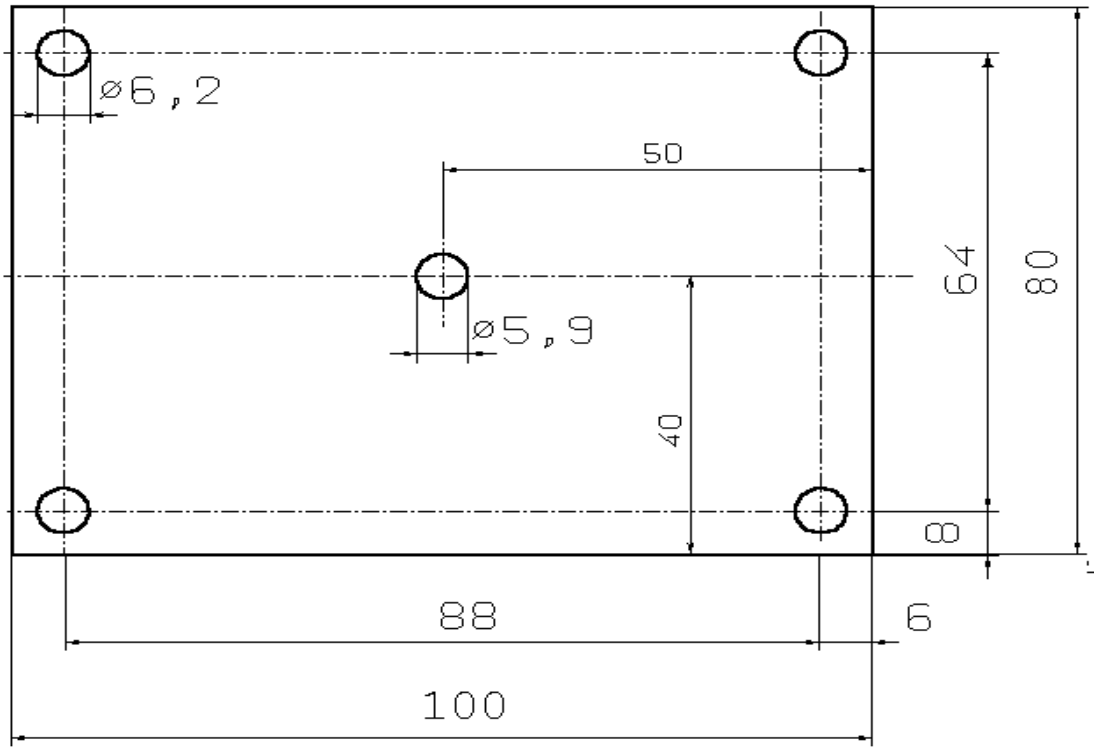
Rotor Diagram – not to scale



Stator Diagram – not to scale

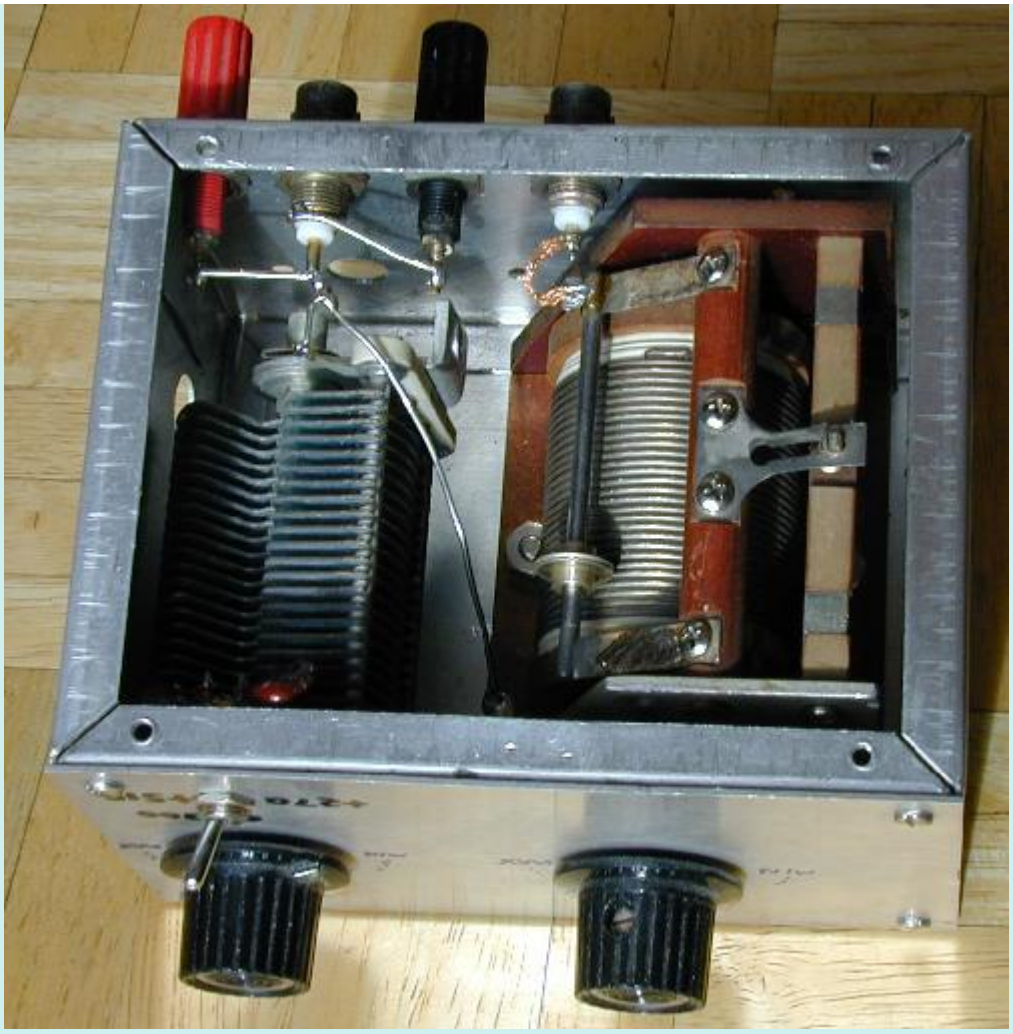


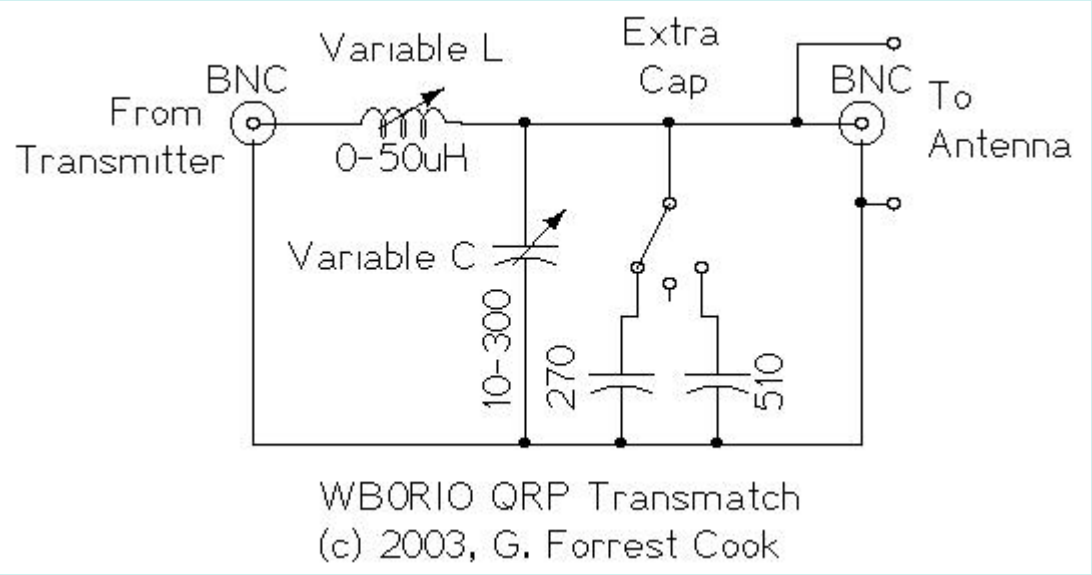
End Plate Diagram – not to scale



QRP Antenna Tuner

(C) 2003, G. Forrest Cook





Introduction

This circuit is for a QRP (low power) antenna tuner (transmatch) for use in the short wave amateur radio bands from 3-30 Mhz. It allows a wide variety of antennas to be connected to a low power transmitter. When the circuit is properly tuned, the maximum transmitter power will be delivered to the antenna. The tuner is normally is used in conjunction with a standing wave ratio (SWR) meter.

This is *not* the most versatile antenna tuner circuit, it will not match any possible load. This tuner *is* efficient and it is very simple to build and use. It is small enough for backpacking applications and is useful for matching many of the types of antennas that one might throw up on a camping trip.

Theory

The purpose of a transmatch is to match the impedance of a transmitter, typically 50 ohms, to an unknown antenna impedance. The circuit consists of a variable series inductor followed by a variable capacitor to ground. Most transmitter outputs consist of a PI low pass network. The transmatch is essentially an adjustable extension of that network. With the parts shown, maximum power through the unit is approximately 50 watts.

Construction

The tuner was built in an aluminum project box. Various holes were drilled in the box for mounting all of the components. When laying out the parts, leave plenty of room around the sides of the components to prevent RF arcing. Be sure to keep both sides of the variable inductor insulated from the box, you may need to use insulated bushings (non-conductive washers) on the inductor's shaft. Use heavy hook-up wire to connect the various components together, I recommend 18 gauge tinned cooper wire, or thicker. Use the shortest wire lengths possible. I used a fairly small box for the components that I had, a larger box would make the project somewhat easier to build.

Use

Connect the transmitter output to an SWR meter, connect the SWR meter output to the input of the antenna tuner, and connect the antenna to the output of the tuner. Antennas with a coaxial feedline should be connected to the BNC output connector. Random wire antennas should be connected to the banana jacks. The transmitter should be connected to a good earth ground at its chassis, the ground can also be connected to the lower (black) banana jack on the tuner.

With coax-fed antennas, the best location for an antenna tuner is where the coax feeds the antenna. Logistically, this may cause the tuner to be in a location that is difficult or impossible for the operator to tune so most people locate the tuner on the transmitter side of the coax.

Set the capacitance switch to 0 (center), adjust the variable capacitor to the mid point. Adjust the inductor to near the minimum inductance. Briefly transmit a CW carrier and observe the SWR reading. If your transmitter has an adjustable output power level, start the adjustment with low power then increase the power and fine-tune when a good match has been found.

Increase the inductance until the minimum reflected power and maximum forward power is observed. Adjust the capacitance for the best SWR reading. It is necessary to go back and forth on the adjustments to find the best match. If the best match is found with the capacitor at the max value, switch in either the 270pf or 510 pf parallel capacitors and re-adjust the variable capacitor and inductor.

Be careful not to leave the transmitter on for too long in the unmatched condition, doing so can damage the transmitter's output transistors. If your transmitter has a variable output power, tune up at low power then re-adjust the circuit at full power. Tube-based transmitters are generally more able to handle large output mis-matches.

Caution: higher power transmitters can generate high voltages within this circuit, don't touch any of the wires when the transmitter is operating. If the roller inductor's adjustment shaft is connected to the inductor's wiring, the shaft should be mounted so that it does not come in contact with the metal box. The set screw on the inductor's knob may be electrically hot during use, you may want to cover it with a drop of plastic glue or silicone after tightening.

Parts

- variable inductor (roller), approximately 0-50 uH

- variable capacitor, 0-300pf or 0-360pf, can be scavanged from an old tube radio
- center off spdt switch
- 270pf, 200V silver mica capacitor
- 510pf, 200V silver mica capacitor
- two BNC connectors (or PL-259, the connector that won WWII, if you prefer)
- two banana jacks
- two insulated plastic knobs
- miscellaneous screws, nuts, and washers
- solid hook-up wire, 18 gauge or similar
- aluminum box, big enough to easily contain all of the components

The variable inductor may be difficult to find, the best places to look are at ham radio swap fests and surplus electronics parts companies. A fixed inductor with switched taps can be substituted. An air-core formed inductor will probably give the best results in this circuit, a toroidal ferrite core inductor will also work, but it may absorb some of the available RF power.

See Wikipedia's Antenna tuner article for more background information and Ulrich L. Rohde's (N1UL) antenna tuner circuit which can tune a wider range of antenna impedances. Ulrich Rohde has also published a more detailed article (zip) in German.

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Design and Implementation of Ambient RF Energy Harvesting Circuits

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***Abstract* - This paper presents design strategies and measurements of two RF energy harvesting circuits working at UHF frequencies (around 480 MHz). Ambient RF sources are considered so for the circuit designs low power levels have been assumed. The circuits were implemented and the RF/DC conversion efficiency was measured.**

***Index Terms* – Matching network, RF energy harvesting, schottky diode, voltage doubler.**

I. INTRODUCTION

In general the energy harvesting denotes all processes through which the energy, coming from alternative sources, as sun and wind, is caught and stored, a particular technique is the RF energy harvesting where the irradiated RF energy is processed. On the other hand, the electrical power levels, generated using this technique, are very low, some milliwatts, but still enough to drive low power devices.

The block diagram of an energy harvesting system is shown in Fig.1, it is composed by the matching network, the multiplier and the DC-load circuit [1]. In this system the matching network permits to maximize the power transfer whereas the voltage multiplier performs the effective conversion from RF to DC. Consequently, the conversion efficiency depends on the type and number of nonlinear devices, which form the multiplier, and is affected by the matching structure.

In addition the characteristics of the RF signal have a direct effect on the RF/DC conversion efficiency, which is characterized by the ratio between the DC power and the RF power.

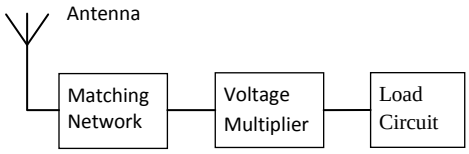

Fig.1. Block diagram of an energy harvesting system.

The RF energy sources can be parted into three general groups: intentional sources, known and unknown ambient sources [2]. This work considers known ambient sources, as antennas that radiate in the UHF frequencies around 480 MHz. In particular, this study is aimed at identifying the potential efficiency of an RF/DC conversion system made with hybrid technology using Schottky diodes and capacitors. Therefore, the voltage multiplier is examined in its simplest form: the voltage doubler. Selected the voltage doubler structures, the study was carried out throw the following steps, firstly, a particular Schottky diode was chosen and its model was implemented on the CAD software used to synthesize the circuits. Secondly, voltage doublers and matching networks were synthesized and realized; successively the achieved circuits were tested in order to verify their correspondence with what was designed. Finally, the real target of this study, measurements to define the RF/DC power conversion were carried out.

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II. SELECTION AND MODELING OF THE DIODE

Schottky diodes can be parted in two basic categories: diodes realized using n or p silicon; the first are characterized by relatively high barrier and low series resistance, consequently they are suitable for detectors and mixers where DC bias is available. On the other hand, the p silicon diodes are characterized by low barrier and high series resistance; these devices were accomplished for application where DC bias can not be used. Therefore, taking into account studies reported in [3], [4] where diode performances are examined the schottky diode HSMS-285C from Agilent were selected. The HSMS-285x family includes zero biased diodes which are optimized for low signal (<-20dBm) applications at frequencies below 1.5GHz.

The diode parameters are reported in the Agilent data sheet in particular the nonlinear behavior of diode is described by the SPICE parameters reported in Table 1[5].

Table 1: SPICE parameters

Parameter	Units	Value
B_V (Vbr)	V	3.8
C_{JO}	pF	0.18
E_G	eV	0.69
I_{BV}	A	3.0E-4
I_S	A	3.0E-6
N	-	1.06
R_S	Ohms	25
P_B (Vj)	V	0.35
P_T (XTI)	-	2
M	-	0.5

The SPICE parameters permit to model the diode chip, however to obtain accurate linear and nonlinear analysis also the package equivalent circuit must be considered [6].

Fig.2 shows the equivalent circuit of the HSMS-285C package (SOT-323) and the component values.

III. VOLTAGE DOUBLERS

The considered voltage doublers are the Greinacher and Delon circuits, the first, shown in Fig. 3, is composed by a Villard cell and a peak detector.

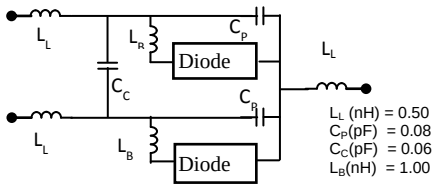

Fig.2. Model of the SOT323 package.

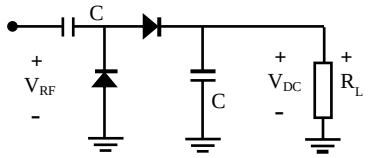

Fig.3. Greinacher circuit.

The effect of the first cell is to shift up the RF waveform so that the negative peaks are clamped to about 0 V. The second cell reduces the ripple providing a DC voltage about equal to the double of RF signal amplitude.

The Delon circuit, shown in the Fig.4, is realized connecting in series two peak detectors, the first detects the positive peak and the second the negative peak, and the total DC voltage is provided across the two capacitors.

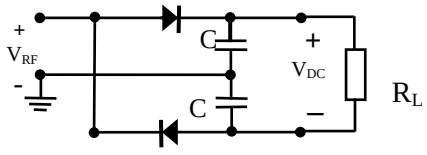

Fig.4 Delon circuit.

In general, the capacitors C do not affect significantly the output voltage. However, the fundamental requirement is to keep their related

impedance low with respect the diode impedance. So capacitors with a value of 100pF are used, they present at the 480MHz frequency a reactance much less than the impedance of a single diode operating at -30dBm. Consequently, the capacitors are short-circuits and the Schottky diodes are put in parallel.

IV. VOLTAGE DOUBLER DESIGN

Chosen the diode, selected the voltage doublers and implemented both on the used CAD software, Advanced Design System (ADS) by Agilent Technoklogies, the design problem consists in the achievement of a matching network between the source at 50 and the voltage doubler input impedance. It is important to note that the diode has an impedance function of the power levels of the RF signal [3],[4]. Therefore, the matching network was synthesized considering a reference power level equal to -30 dBm. In particular, to reduce losses due to passive components, the simplest circuit solutions have been chosen, namely the L networks.

The Fig.5 shows the achieved Greinacher doubler, where the matching network is constituted by the components C_1=0.8pF and L_1=56nH. The element HSMS285C represents the pair of diodes and the package model.

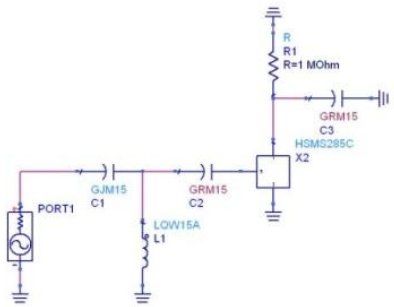

Fig.5. Simplified scheme of Greinacher doubler.

The resulting input impedance of the doubler, without and with matching network, is shown in Fig.6.

The Fig 7 shows the Delon doubler, where in the L matching network C_1=0.9pF and L_1=68nH result. Fig. 8 describes the doubler input impedances without and with the L network.

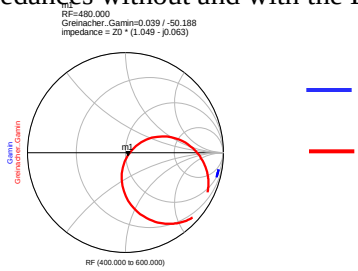

Fig.6 Greinacher doubler input impedance.

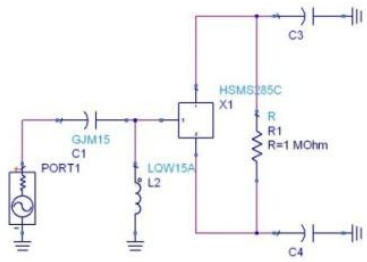

Fig.7 Simplified scheme of Delon doubler.

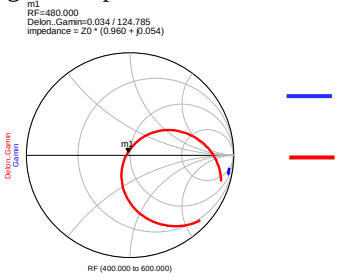

Fig.8 Delon doubler input impedance.

For simplicity in the schemes shown in Fig.5 and Fig. 7 the microstrip connections are not reported. However, in the real design, microstrips, vias have been taken into account. Moreover, to ensure a good match between simulations and measurements the structure discontinuities have been characterized using the Momentum, which is electromagnetic simulator of ADS.

For the two doubler achievements is assumed the use of FR4 (e_r = 4.35, substrate thickness =1.6 mm, conductivity = 5.8 e7 S/m, conductor thickness = 35 μm, tan δ = 0.018).

V. VOLTAGE DOUBLER REALIZATION

In order to perform the desired measures the two voltage doublers have been realized and the obtained circuit photos are shown in Fig. 9 and Fig. 10.

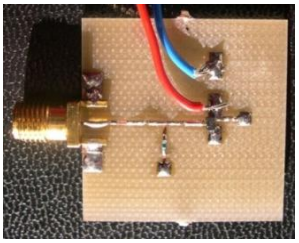

Fig.9. Greinacher doubler photo.

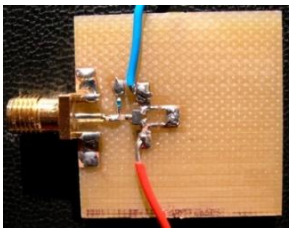

Fig.10. Delon doubler photo.

Measurements by using Anritsu 37397D VNA were done to verify the fitting between designs and produced circuits, and S_{11} measures and simulations were compared for different signal power levels. In particular Fig.11 and Fig.12 show the return loss, obtained assuming a power level equal to -30 dBm.

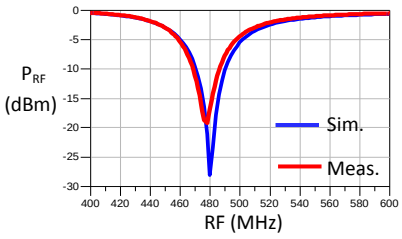

Fig.11. Return loss of the Greinacher doubler.

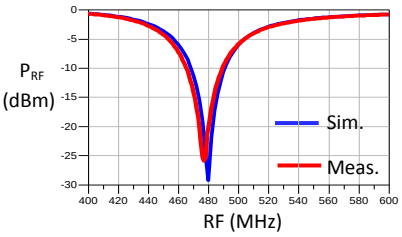

Fig.12. Return loss of the Delon doubler.

VI. MEASUREMENTS

Verified the correspondence between design and implementation for the voltage doublers, measurements were carried out to obtain the DC voltage values resulting from the RF input. This result was achieved using as RF source the Agilent E4438C vector signal generator. Several measurements were made for different loads and RF power levels; Fig.13 and Fig.14 show the obtained voltage values for RF power levels equal to -40, -30 and -20 dBm for the two selected configurations.

The best performance are obtained around 480MHz and, for a load equal 10MΩ, an input RF power of 1μW produces about 70 mV. Obviously, for lower load values, the results decrease, so at 100KΩ and 10KΩ the DC voltage values are around 55 mV and 25 mV for an input RF power equal to 1μW.

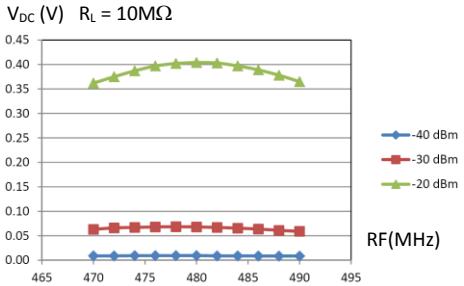

Fig.13. Voltage values by Greinacher doubler.

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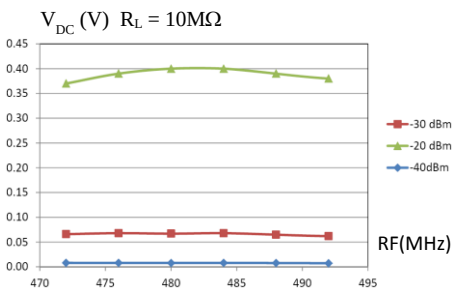

Fig.14. Voltage values by Delon doubler.

The collected measurements were processed in order to achieve the RF/DC power conversion efficiency. The Fig.15 shows the conversion efficiency values obtained for different power levels and different loads. So at a 1μW the greatest conversion efficiency value is about 6% achieved with a load of 10KΩ.

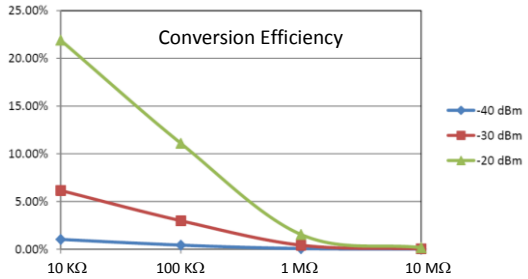

Fig.15. RF/DC conversion efficiency by Greinacher doubler.

VII CONCLUSION

This work was aimed to acquire knowledge on the RF/DC conversion efficiency values which can be achieved by hybrid technology, using Schottky diodes and capacitors. UHF TV broadcasting frequencies around 480 MHz have been taken into account. The desired result was performed, firstly designing and realizing two different RF/DC conversion systems and then measuring and processing the collected data. In particular, for a load equal 10 MΩ with an input RF power of 1μW a DC voltage equal 70 mV was achieved.

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[4] Agilent technologies application note note 1088, "Designing the virtual battery ".

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[6] Agilent technologies application note 1124, "Linear Models for diode surface mount packages ".

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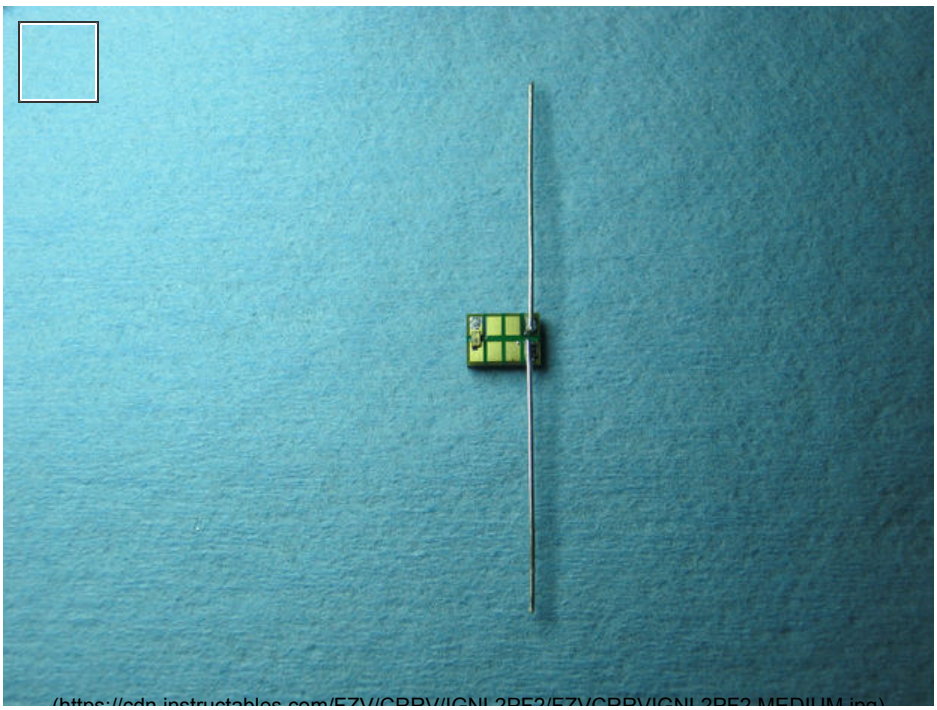

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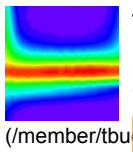

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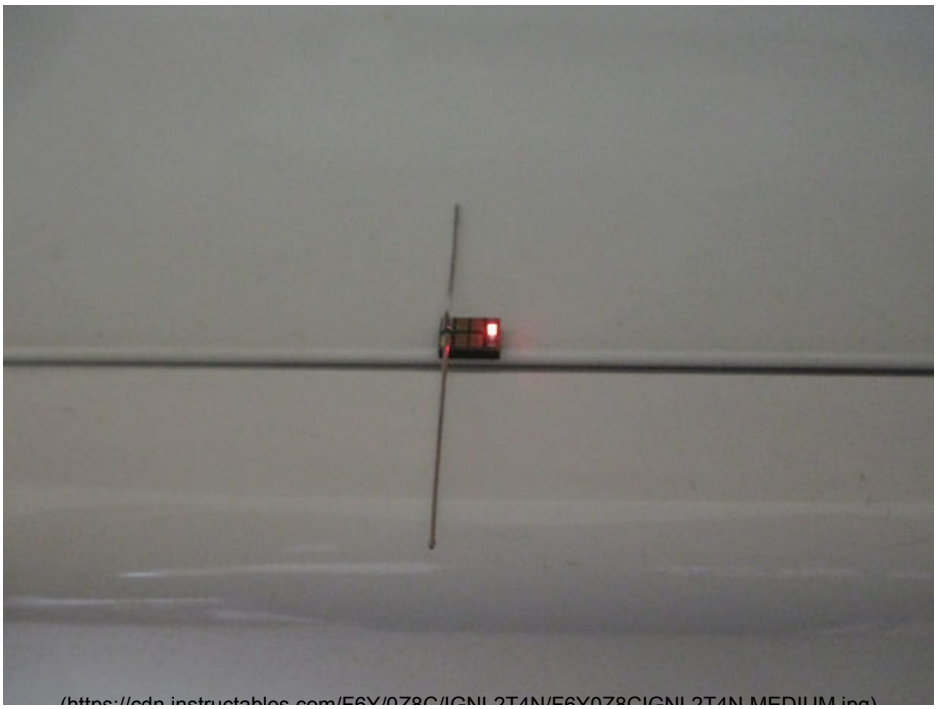

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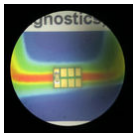
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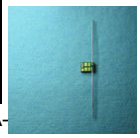
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Description

This article describes how to build a wireless harvesting device that will capture stray 2.5GHz energy from a microwave oven and convert it into a useable voltage that will light up a red light emitting diode (LED). Hobbyists can modify the antenna with longer wires and try to pick up other signals such as AM/FM radio, mobile phone signals and other wireless energy. The LED's light will be visible when the module picks up 1 milliwatt-10 milliwatts of microwave power. Usually the most power leaks out of the door seam of a microwave oven.

What you need:

1 RFD102A (RF-DC Converter) module available at

http://www.rfdiagnostics.com/product/rfd102a/

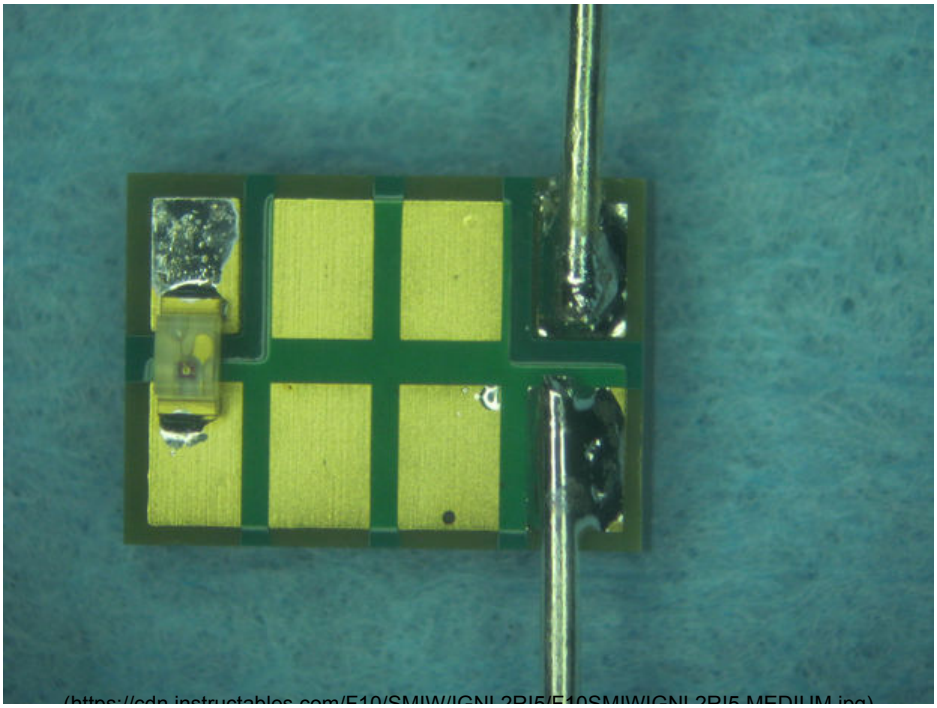
1 ¼ watt leaded resistor (any value) or two wires that are 1 1/8 inches long (28.6mm)

1 Kingbright Surface mount red LED (part number: APT1608EC) available at Digikey/Mouser/Newark

Fine tipped soldering iron

Solder paste or Fine Solder

Step 1: Assembly Instructions

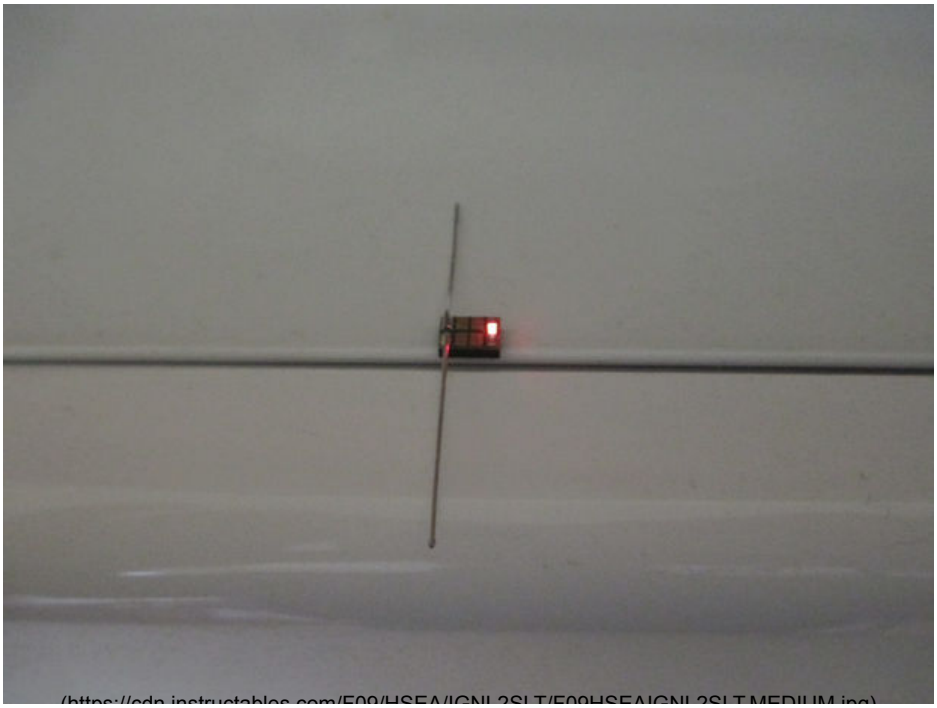


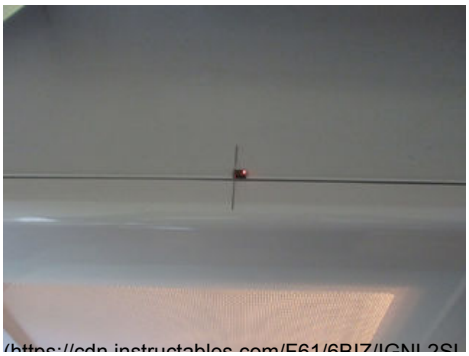
Cut the resistor wires off next to the resistor. These are just the right size at 1 1/8” long for a 2.5GHz dipole. Throw away the resistor and keep the wires.

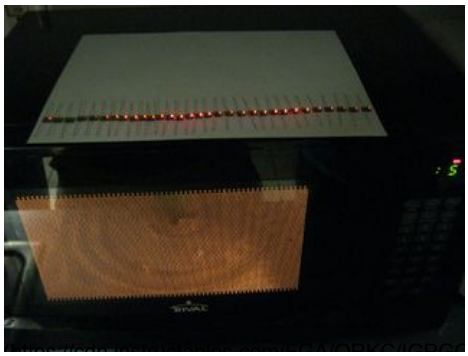
Put solder paste on the module at pins 1 & 8 and at pins 4 and 5.Place the wires on pins 4 and 5 and solder carefully using tweezers to hold the wires (it will burn you otherwise). Solder at the lowest soldering temperature possible to avoid damaging the module. If the iron is too hot then you may damage the internal connections inside the module. Use a minimum of time for soldering (<10secs). The wires work as a dipole antenna to collect the 2.5GHz energy into the RF (Radio Frequency) Input of the module.

Place the LED with the anode (positive side) onto pin 1 and the cathode (negative side) on pin 8 and solder carefully. For those not familiar with LEDs, the triangle symbol of the diode should point to the ground pin of the module (pin 8). Your final microwave harvester should look like figure 2

Step 2: Test on Your Microwave Oven







Now put some water in a mug and put it in the microwave oven and cook on high for 2 minutes. While the microwave is running, hold the module and move the module antenna around to find a hot spot on the microwave. Tape the module on the ground wire to the microwave to keep it in place. If all is well then you will see the LED light up and dim as the electric fields change while the mug rotates in the microwave oven. The above figures show the microwave energy harvester in action. Experiment with antenna shapes and sizes. If you go too big or too small with the antenna then less energy will be collected. You can also put the assembled module near an 802.11b/g/n wireless LAN router and watch it flicker as the router transmits and receives 2.4-2.5GHz energy. Enjoy!

Other Project Ideas:

- Harvest AM/FM Radio Signals

FAQ

- What is the maximum current this module can provide? Ans: Typical current is ~0.5-5mA. Max current is 18mA.

- What is the efficiency of converting RF energy to a DC voltage? Ans: If the input power to the module is around 10-50mW then the RFD102A can achieve >50% efficiency if the resistance seen by the DC output is ~10kOhm.

- I don't have a soldering iron and am all thumbs. Where can I get a detector like this already built up? Ans. The RFD102A-DET is available http://www.rfdiagnostics.com/product/microwave-oven-window-energy-

harvesting-kit/

- What is the maximum output voltage this module can produce? Ans: With 0.5W input power the module can produce up to 37V at 915MHz.

- Can I charge my mobile phone with this? Ans: It is possible with 4 RFD102A modules and antennas to produce enough current to charge a mobile phone. The hard part is the wireless source would need to be so powerful that you wouldn't want to be in the same room during charging. You may hurt your eyes by cooking them.

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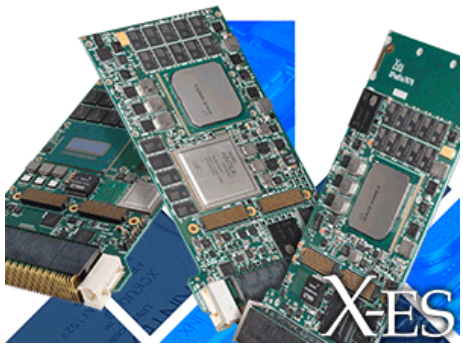

Comments

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ShamarY (/member/ShamarY/) 2016-01-09 Reply

is it possible that instead of connecting the led a speaker is put in its place and a phone can actually be listened in on, or the radio?

Microwave Energy Detector & Wireless Energy Harvesting by tbudka (/member/tbudka/) in electronics (/technology/electronics/)

Download (/id/Microwave-Oven-Energy-HarvestingDetector/) 2 Steps

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utg001 (/member/utg001/) 2015-11-13 Reply

I had an explosion of ideas from this project, I had to make this account :p

1> you seem to use specific dimensions/material of wire to tune it to 2.5 GHz, whats going on here? Can I use a sheet of metal? Can I tune it to some other frequency on this module? If so, how?

2> What is the ideal output from this model you made under ideal conditions (V & I)? Is it versatile enough to be used as a small power source for something if I join enough of them in series and parallel? Say 1000 of them in 10x10 series and parallel connection harnessing a constant strong signal, can the module support a tiny load (bit more than LED)?

I'm sorry if it sounds too stupid, but I'm only wanting to know a bit more fine details before I head to the drawing board :)

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tbudka (/member/tbudka/) ▸ utg001 (/member/utg001/) 2015-11-14 Reply

1. Dipole antennas are usually sized to be 1/2 of a wavelength so each wire in this dipole is 1/4 of a wavelength. At 2.5GHz the 1/4 wavelength (c/f) = 0.25 * (3.0E10cm/sec)/(2.5E9 Hz) = 3cm. The leaded resistor wires are 2.86 cm (1 1/8 inch) long so that checks out. Sheet metal could be used for a bow-tie antenna or a cross pattern to pick up the other polarization. Size does matter. A larger dipole will pick up lower frequencies. The RFD102A module is broadband and limited by the frequency response of the antenna.

2. I posted a photo at the end of an array of 30 detectors picking up the energy. Your idea of putting a 10x10 array is a good one to scale up the power harvested. Kind of like having a larger solar panel. That would be a cool project to make.

Thanks for your questions/comments.

ke8bg (/member/ke8bg/) 2015-11-14 Reply

From the FDA website on leakage..

Use to work on microwave ovens.. We had to leakage check each one before they went back into service. Most likely leakage is around the door. A door out of adjustment can spew out all sorts of radiation. Redundant door safety switches are there for a reason. I've been 'dosed' a couple of times by a technician who had left off the brass Magnetron gasket and couldn't figure out why the control board would shut down right after he pushed the start button. 500 watts at 2.4 ghz will overwelm a CPU chip. Be careful. 2.4ghz can cook you too!!!

BTW as stated below the power level drops off 'inverse square of the distance'.

"A Federal standard limits the amount of microwaves that can leak from an oven throughout its lifetime to 5 milliwatts (mW) of microwave radiation per square centimeter at approximately 2 inches from the oven surface. This limit is far below the level known to harm people. Microwave energy also decreases dramatically as you move away from the source of radiation. A measurement made 20 inches from an oven would be approximately one one-hundredth of the value measured at 2 inches."

tbudka (/member/tbudka/) ▸ ke8bg (/member/ke8bg/) 2015-11-14 Reply

Thanks for this information and it's nice to meet someone familiar with microwave leakage/exposure. The light from the detector definitely gets dimmer very quickly as I pull it away from the microwave door. 5mW is very small and is only seen with this detector close to the microwave oven (<2inches). This is why I think we can only harvest energy close to the microwave rather than at the kitchen table. I would need to add an amplifier to pick up the weak microwave energy from the microwave if I were >5ft away.

MarcoG6 (/member/MarcoG6/) 2015-11-08 Reply

there was a kickstarter that made a phone case to turn the stray signals from ur phone to charge the phone they claimed to get 3x the battery life in a iphone 5 but the case was like 100$ maby u can make a chaeper one

Dashing Rainbow Dash (/member/Dashing+Rainbow+Dash/) ▸ MarcoG6 (/member/MarcoG6/) 2015-11-13 Reply

tbucka is right, it just doesn't seem... legit.

Dashing Rainbow Dash (/member/Dashing+Rainbow+Dash/) ▸ Dashing Rainbow Dash (/member/Dashing+Rainbow+Dash/) 2015-11-13 Reply

*tbudka sorry

tbudka (/member/tbudka/) ▸ MarcoG6 (/member/MarcoG6/) 2015-11-09 Reply

That claim doesn't seem realistic. If the charger is collecting a lot of power from the phone then it's probably hurting its transmission/reception. The RFD102A module is very sensitive to low input powers but didn't get much power from my phone with a log periodic antenna. At 0.04mW/915MHz the module produces 0.5V output into a voltmeter. When I was testing the RFD102A-TB (TB=test board) my phone while on a call to my answering machine, I got up to 4V output from the module but that is into a high impedance. I think my next instructable will document this tests with a phone and a base station. Thanks for your comment.

MichiganDave (/member/MichiganDave/) 2015-11-07 Reply

A BIG Thanks for sharing and a small question for you: If you can harvest this energy can you save it? I am imagining doing this in order to save the "free" energy until its useful. Will it take a thousand years to power a flashlight, or merely ten?

tbudka (/member/tbudka/) ▸ MichiganDave (/member/MichiganDave/) Reply 2015-11-08

To answer your question about charging a flashlight, yes it is possible to charge a flashlight in a reasonable time with your microwave oven if you have ~10-20 of these detectors to generate 50-100mA to charge up the batteries. Please see my last photo with 30 detectors on a microwave all lighting up. Then you need to make sure the oven is on a lot (like at Subway). Unfortunately you need to be so close the microwave to get this power. So IMHO, if you are going to try to get free energy, you need a good source in mind. Stray cell phone and WiFi power aren't high enough. An array of antennas pointing at a base station may give you some small amount of power but it won't be cost effective.

tbudka (/member/tbudka/) ▸ MichiganDave (/member/MichiganDave/) Reply 2015-11-07

Yes. You can save the energy but not indefinitely. There are always small leakage paths to rob you of this hard fought free energy. The RFD102A has 60kOhms resistance to GND so if you had been able to charge a supercap (1F) to 5V, it would take 60000secs (1000mins or 16.67hrs) to discharge the capacitor by 36.8% of its energy (RC time constant). Supercaps also have their own leakage resistance. So you need to make sure you understand your microwave source and the leakage resistances of your components. TinkerJim has a great instructable on a Solar Engine that is very useful with this module. I am planning on posting about an improved module with TinkerJim's design. Another great question. Thanks for your interest.

turbiny (/member/turbiny/) 2015-11-06 Reply

Is it possible to recharge a phone via wifi router?

tbudka (/member/tbudka/) ▸ turbiny (/member/turbiny/) 2015-11-06 Reply

No. Most WiFi routers are very low output power. If you have a WiFi Booster Amplifier (2W) then there is a chance to get some power from this system to charge your phone but I wouldn't want to stay in that room with all that power. Most WiFi signals are <0.05W coming out of the router. This detector will light up and flicker close to an 802.11b/g router (2.5GHz) but that's only a few milliamps of current at 1.5V. Sorry about this but I agree with Dave Jone's eeVblog #55 that you won't be able to charge your cell phone from WiFi unless the transmitted power gets raised. Great question.

whyjbe (/member/whyjbe/) 2015-11-06 Reply

Thanks tbudka for the tips. It's good to know that the power level drops quickly with distance. My microwave is still 'new' since I do not use it.

tbudka (/member/tbudka/) ▸ whyjbe (/member/whyjbe/) 2015-11-06 Reply

There is really not much to worry about with microwave ovens, cell phones and base stations for health issues. I tested this module with a broadband antenna around a base station and I got a very low voltage out (1.5V max) which would mean there is less than 0.0005W of power right next to a base station. The cell phone companies do a great job keeping that power low around the base stations. Also since the wavelengths are large (inches/cm) there can't be much if any damage at the cell level (microns). IMHO I don't see much of a chance for low power (<10mW) microwaves to cause any health problems. Just don't stare at the microwave popcorn right next to the window... :)

whyjbe (/member/whyjbe/) 2015-11-06 Reply

Thanks tbudka for the tips. It's good to know that the power level drops quickly with distance. My microwave is still 'new' since I do not use it.

whyjbe (/member/whyjbe/) 2015-11-06 Reply

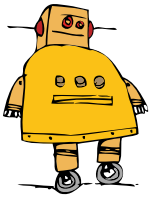
So is this another way of saying that this module transmutes the harmful EMF (electromagnetic frequency) waves coming off the microwave since it converts it to electrical power? If so that's awesome!

tbudka (/member/tbudka/) ▸ whyjbe (/member/whyjbe/) 2015-11-06 Reply

This detector helps give the average person a way to "see" the levels of microwave energy leaking out. The power level drops extremely quickly with distance from your microwave so usually there is no need to worry about it. As you move this detector more than a few inches from most microwave ovens you won't see any light from the LED. This indicates that the power is less than 1mW which is tiny for exposure. For microwave

ovens, the main danger is for your eyeballs where heating could cause cataracts. I don't believe there are any other harmful effects than that from your microwave at powers <10mW. If you have a loose door then I would recommend getting a new microwave. Please don't put your eyes close to the window or door seam or side vent of your microwave.

More Comments



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Wireless device converts ‘lost’ microwave energy into electric power

November 8, 2013

Using inexpensive materials configured and tuned to capture microwave signals, researchers at Duke University’s Pratt School of Engineering have designed a power-harvesting device with efficiency similar to that of modern solar panels.

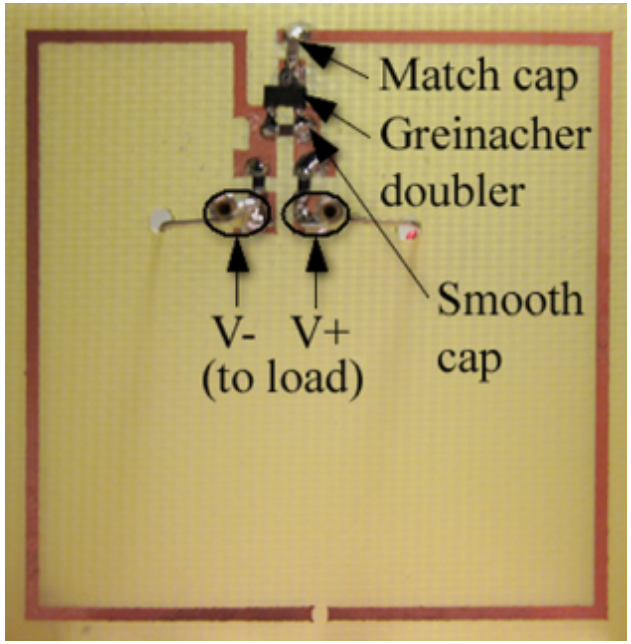
The device wirelessly converts a microwave signal to direct current voltage that is capable of recharging a cell phone battery or other small electronic device.

It operates on a principle similar to that of solar panels, which convert light energy into electrical current. But this versatile energy harvester could be tuned to harvest the signal from other energy sources, including Wi-Fi signals, satellite signals, or even sound signals, the researchers say.

The key to the power harvester lies in its application of metamaterials, engineered structures that can capture various forms of wave energy and tune them for useful applications.

Undergraduate engineering student Allen Hawkes, working with graduate student Alexander Katko and lead investigator Steven Cummer, professor of electrical and computer engineering, designed an electrical circuit capable of harvesting microwaves.

They used a series of five fiberglass and copper energy conductors wired together as an array on a circuit board to convert microwaves with an RF-to-DC conversion efficiency of 37 percent.



Prototype power harvester resonant at 900MHz (a GSM cell-phone frequency) (credit: Allen M. Hawkes at al./*Applied Physics Letters*)

The efficiency of solar cells

That efficiency is comparable to what is achieved in solar cells, according to Hawkes.

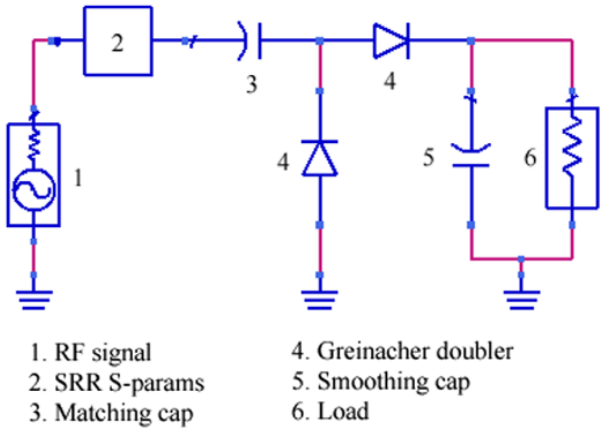
“It’s possible to use this design for a lot of different frequencies and types of energy, including vibration and sound energy harvesting. Until now, a lot of work with metamaterials has been theoretical. We are showing that with a little work, these materials can be useful for consumer applications.”

For instance, a coating could be applied to the ceiling of a room to redirect and recover a Wi-Fi signal that would otherwise be lost, Katko said. Another application could be to improve the energy efficiency of appliances by wirelessly recovering power that is now lost during use.

"The properties of metamaterials allow for design flexibility not possible with ordinary devices like antennas," said Katko. "When traditional antennas are close to each other in space they talk to each other and interfere with each other's operation. The design process used to create our metamaterial array takes these effects into account, allowing the cells to work together."

Wireless recharging for cell phones

With additional modifications, the researchers said the power-harvesting metamaterial could potentially be built into a cell phone, allowing the phone to recharge wirelessly while not in use. This feature could, in principle, allow people living in locations without ready access to a conventional power outlet to harvest energy from a nearby cell phone tower instead.



Schematic of power harvester (credit: Allen M. Hawkes at al./*Applied Physics Letters*)

"Our work demonstrates a simple and inexpensive approach to electromagnetic power harvesting," said Cummer. "The beauty of the design is that the basic building blocks are self-contained and additive. One can simply assemble more blocks to increase the scavenged power."

For example, a series of power-harvesting blocks could be assembled to capture the signal from a known set of satellites passing overhead, the researchers explained. The small amount of energy generated from these signals might power a sensor network in a remote location such as a mountaintop or desert, allowing data collection for a long-term study that takes infrequent measurements.

The research was supported by a Multidisciplinary University Research Initiative from the Army Research Office.

Abstract of *Applied Physics Letters* paper

We present the design and experimental implementation of a power harvesting metamaterial. A maximum of 36.8% of the incident power from a 900 MHz signal is experimentally rectified by an array of metamaterial unit cells. We demonstrate that the maximum harvested power occurs for a resistive load close to 70 Ω in both simulation and experiment. The power harvesting metamaterial is an example of a functional metamaterial that may be suitable for a wide variety of applications that require power delivery to any active components integrated into the metamaterial.

References:

- Allen M. Hawkes, Alexander R. Katko, Steven A. Cummer, A microwave metamaterial with integrated power harvesting functionality, *Applied Physics Letters*, 2013, DOI: 10.1063/1.4824473

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CRYSTAL SETS 1
Introduction
What - Why - How

Crystal Sets & other pages:

- Build Your Own Crystal Set (Pt 2)
- Spider's Web Crystal Set (Pt 3)
- Crystal Set By Kenneth Rankin (Pt 4)
- Experimental Crystal Set (Pt 5)
- TRF Radios
- Riding On A Wave
- Loop Aerials and ATU's
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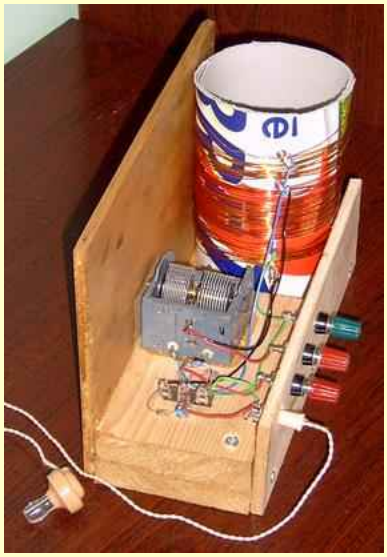
An old fashioned style detector that is housed in a glass tube. The point of the cat's whisker touches the crystal (galena) on the left & is adjusted with the black handle.



CRYSTAL SETS PART 1 - INTRODUCTION

What They Are, Why They Came Into Being and How YOU Can Build Your Own CRYSTAL SET!

What are Crystal Sets? I am glad you asked! It is a fascinating subject dating back to 1920 when Marconi first started experimenting with radio broadcasting from his station 2MT in Chelmsford.

The earliest listeners to Marconi's experimental radio programmes would have mainly used the most basic of radio sets called a Crystal Set. Companies such as the W.G Pye began producing early radio sets in 1921 which used electronic components called 'valves', the forerunner to today's transistors and silicon chips). These glass valves could be arranged to amplify the weak radio signal to make them louder - enough to operate a loudspeaker. However these radios were expensive and needed electricity or bulky lead acid batteries to operate.

THE MAGIC OF A CRYSTAL SET

The crystal set was popular during the 1920's and 30's because it was cheap to buy, much cheaper if you built one yourself, perhaps only a few shillings. Additionally a Crystal Set did not require electricity or expensive batteries, however it could only provide enough volume for headphone listening, and if it was situated in an area of low signal strength the listener would also require a very quiet room as the volume would be low. However In the 1920's hearing voices and music from a station miles away with no wires (hence the term Wireless) really was a magical experience! Listeners in the early twenties were also tinkerers and experimenters, trying different designs of crystal set in an attempt to get the loudest and clearest reception.

Since crystal sets have no batteries and no mains power, they rely entirely from the electrical energy developed between the aerial an earth connections - producing sounds as if by magic.

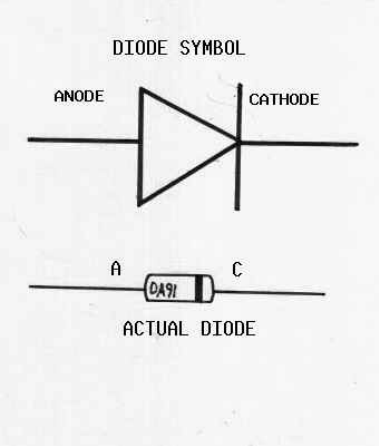
The crystal set had to be carefully tuned into the station by making adjustments to a tuning coil and condenser, these early crystal sets had the added complication of using a 'Cats Whisker' as the detector. The detector converted the radio waves received from the radio station into an audio wave that could be heard in the headphones. The Cats Whisker was a fine wire that rested onto a piece of galena, the crystal, and had to be very finely adjusted to obtain the loudest and clearest sound. Once the 'sweet spot' was found it was important not to move it, it would be very frustrating if someone bumped into the table and dislodged it!

To work at all a crystal set requires a very large aerial, but this is nothing more than a long piece of wire, perhaps 20 to 50 yards of insulated copper wire, hung outside around a garden between poles or trees, or even around the loft space. For the aerial to be effective it needs to be balanced with an earth. An earth is simply a copper rod or pipe about one yard long driven into some soft damp earth with a mallet. At the top of the pipe is fitted a jubilee or hose-clip to which is attached a length of insulated copper wire which is then fed into the house, along with the aerial wire, to the room where the crystal set will be operated and connected to the set's terminals.

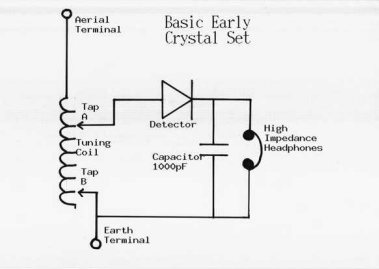
It is quite possible to build your own crystal set. In its very simplest early form a crystal set consisted of a coil of wire, something called a 'detector' and a pair of very special headphones.

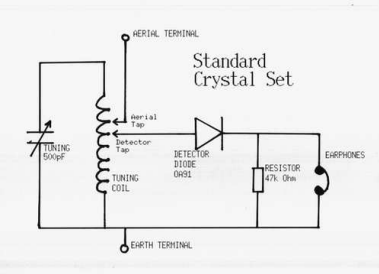
THE CONSTITUENT PARTS

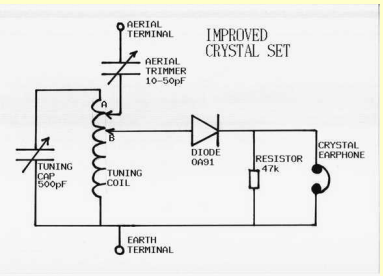
Above: A REAL Crystal Set
You Can Build One Too!
HERE



An illustration of a modern diode which is often encapsulated in glass and is about 7mm long. Above is the electronic symbol for a diode.







The coil of wire is called a Tuning Coil, and the number of turns of wire on the coil determines the wavelength (frequency) that the set is tuned into. Tuning to different stations can be accomplished by varying the number of turms on the coil, or more easily by having many different tapping points on the coil so that adjustments can be made.

To make tuning easier a component called a tuning capacitor can be included in the circuit. In very early sets a tuning capacitor was not always included to keep costs down, or due to their being difficult to obtain.

The Detector converts the radio wave received into an electrical wave that is suitable for the headphones to, in turn, convert into sound waves that can be heard by the human ear. In the very early days of the crystal set the detector consisted of a holder containing a piece of galena crystal that had a very thin and springy wire placed on its surface that had to be very delicately adjusted to find the sweet spot where the radio station could be heard. This was commonly referred to as a "Cat's Whisker". Modern detectors are called diodes and are more efficient than early detectors and cats whiskers. Diodes are still quite readily available and inexpensive. Part numbers for modern diodes include OA90, OA91 and IN34. Diode part numbers that are perhaps now more difficult to obtain include OA47 and OA81.

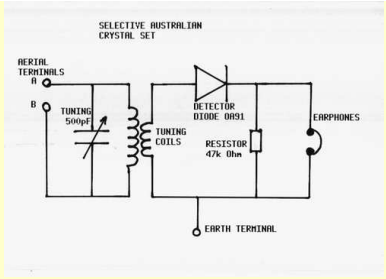
The headphones have to be of a very special type called high impedance headphones. Because there is no additional power source in a crystal set the current generated in the circuit is tiny - minuscule in fact. Ordinary low impedance headphones, such as Walkman headphones, would present a virtual short circuit to the crystal set allowing the tiny signals to drain away to earth & consequently producing no sound - not very useful!

High impedance headphones, on the other hand, reduce or impede the flow of current down to earth, in effect saving the tiny signals to produce sounds from the headphones that we can then hear.

There is a problem however, these high impedance headphones that were so readily available in the 1920's and 1930's are to so easy to obtain today, but some specialist vintage radio outlets still stock them though the price can be quite high. They can still be seen in museums of course. All is not lost though, today we can obtain a special earphone called, appropriately, a crystal earpiece very easily and far more cheaply than. An electronic component, called a resistor, must be connected accross the crystal earphone to allow a path for DC current to get to earth. The value of the resistor is usually 47,000 Ohms and without it a crystal earphone tends to block DC current and as a consequence the sound will be very quiet and distorted.

Aerial and Earth: For a crystal set work it needs an efficient aerial and earth. Since a crystal radio has no power - no batteries or mains electricity - it relies entirely on the radio wave energy sent out from the radio station's transmitter and collected by the aerial to work. The aerial simply consists of a length of wire, but it necessarily has to be quite long, usually in the order of 10 to 20 meters. For the aerial to be effective the crystal radio set also has to be connected to a good earth point. A good earth often consists of a 3 or 4 foot copper stake driven into the ground, but sometimes a water pipe can be used to reasonable effect. (Safety: Never use the earth pin of a household mains plug)

The circuit diagram (schematic) for a crystal set is shown below. Practical designs are shown on the following pages.



No AM radio stations or transmitters in your locality or country?

Has your local medium wave broadcast station closed or been moved to VHF/FM or Digital? Don't worry! You can still build and experiment with crystal sets and TRF radios by also building or buying a simple low power AM transmitter. So, not only can you use your crystal sets but you can also run your own radio station that can be heard in and around your home - playing the music or programmes that you want to hear!

Spitfire & Metzo Complete high quality ready built medium wave AM Transmitters from Vintage Components:
http://www.vcomp.co.uk/index.htm

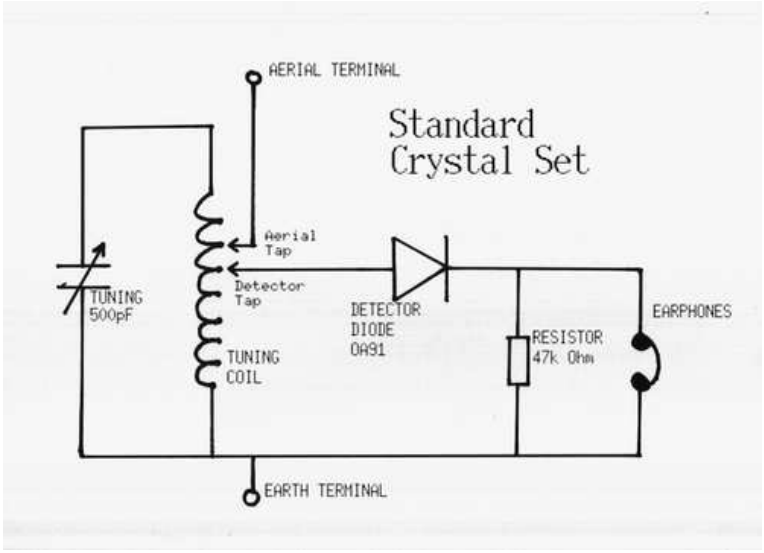
Choice of Spitfire and Metzo transmitters:
Spitfire:
http://www.vcomp.co.uk/spitfire/spitfire.htm

Metzo : AM transmitter including audio compression circuits:
http://www.vcomp.co.uk/metzo/metzo.htm

SSTRAN AMT3000 : Superb high fidelity medium wave AM transmitter kits from SSTRAN. Versions available for 10kHz spacing in the Americas (AMT3000 or AMT3000-SM) and 9kHz spacing in Europe and other areas (AMT3000-9 and AMT3000-9SM). Superb audio quality and a great and well designed little kit to build:
http://www.sstran.com /pages/products.html

http://www.sstran.com/

AM88 LP : AM transmitter kit from North County Radio.
http://www.northcountryradio.com /Kitpages/am88.htm



Above: The circuit diagram for the standard Crystal Radio

Above: A commercially made crystal set from around 1940

There is no variable tuning capacitor (condenser) in the above design. The capacitor is a fixed component and tuning is achieved by varying the coil inductance with the rotating switch. The detector is a galena crystal, seen at the top of the panel. A pair of high impedance magnetic headphones can just be seen to the right of this shot. (photo sent in by our reader Krysatec in Czechoslovakia - thanks!)

EXCELLENT RESULTS

Almost certainly when building a crystal set today a crystal earphone will need to be used. A crystal earphone is not quite as good as the old high impedance headphones as it is not quite as sensitive and therefore not quite so loud, however today's transmitters are much more powerful than those of the 1920's and so really excellent results are very often obtained - just don't expect to hear everything that you might on a top quality portable radio!

Crystal sets are fascinating because they are so simple to build, and because there is no power required they remain magical to this day. I have built several over the years, with differing designs, and if you'd like to build one yourself, it's easy, and I have described some circuits in part 2 with some more experimental ideas in part 5.

I have a crystal set that receives BBC Radio 4 on long wave and 3 local stations and 3 national stations on medium wave and another that also receives shortwave

stations. Not bad for radios with no battery or other power - and such as set could be very useful in an emergency.

More About Diodes - unconventional and new ideas (by Felix Scerri)

While messing around with diodes for crystal sets recently, I tried something as a theory that actually works very well in practice, and with one or two actual advantages.

Ordinarily the ubiquitous 1N914 or IN4148 silicon switching diodes are pretty hopeless as RF detectors unless RF signals are *very* strong! However there is a simple way of making these diodes work very well as RF detectors, by the addition of adjustable voltage bias. I have used adjustable voltage bias with diodes before with very good results, but using bias with IN914 or 1N4148's is a relatively new idea!

Interestingly enough with voltage bias correctly adjusted, 'weak signal' performance and sound quality is actually very good, with that typically 'crisp' silicon diode 'sound'. I use a simple bias network with a 1.5 v AA battery and a 1 Megohm potentiometer applied to the diode directly with the incoming RF fed into the diode through a 1 uf plastic film capacitor (as shown). The use of adjustable bias seems to help most with 'repositioning' the position of the knee in the diodes curve and weak signal performance is excellent, easily as good as any 'proper' germanium diode.

Another 'tweak' that is actually quite beneficial is 'paralleling' several diodes together. This doesn't work with ordinary germanium diodes but definitely works with BAT46 'germanium equivalent' diodes and the effect of three of these diodes together is to make a better 'single' diode. The turn on voltage is noticeably reduced and the 'square law; region of detection is definitely improved and so is the rectified audio quality. These diodes are relatively inexpensive (at least in this country), so it is worth trying. 73 Felix vk4fuq 080214.

Part 2
Practical Designs To
BUILD YOUR OWN CRYSTAL SET!

Having difficulty in finding components? I have added some ideas for component sources here.
Sources For Older Components

Build Your Own Crystal Set (Part 2) | **Spider's Web Crystal Set (Part 3)**

Crystal Set By Kenneth Rankin (Part 4) | **Experimental Crystal Set (Part 5)**

LINKS

Components:

BOWOOD ELECTRONICS - A friendly, helpful and very speedy source for your electronic components at prices that won't frighten your wallet! - The MK484 IC (the ZN414 replacement) is available from here.

VINTAGE COMPONENTS - A great resource for crystal sets, components, valve radio kits and medium wave AM transmitters!

6V6 - Electronic Nostalgia and Vintage Components

1N34A.com - crystal radio parts, including litz wire is http://1n34a.com

More Links To Electronic Component Suppliers here

Crystal Sets and other interesting radio related topics:

ALSO: BUILD A RADIO IN A MATCHBOX!

Click To Find Out More!

- **Amateur Radio by MØMTJ**

OZ CRYSTAL RADIOS by Austin Hellier

Simple crystal set receivers used by soldiers during the war and by prisoners of war (P.O.W.'s):
THE FOXHOLE and P.O.W RADIOS

Dave Schmarder's Crystal Set Pages
http://www.makearadio.com/crystal/index.php http://www.makearadio.com

"Heart of England Crystal Radio Club" based in Birmingham, West Midlands - http://crystalradioclub.co.uk

Stay Tuned http://www.crystalradio.net/
Designs http://www.crystalradio.net/crystalplans

The Tuggle Two Crystal Set
http://www.crystalradio.net/crystalsets/miketuggle/index.shtml

Jim's Crystal Radio Page http://www.hobbytech.com/crystalradio/crystalradio.htm

VE7SL Crystal Radio DXing http://members.shaw.ca/ve7sl/crystal.html

Gollum's Crystal Receiver World http://www.oldradioworld.de/gollum/

High Fidelity AM/MW Broadcast Band Reception by Felix Scerri
http://sound.westhost.com/articles/am-radio.htm

Crystal Set by G3XBM http://homepage.ntlworld.com/lapthorn/xtal.htm

Experimental Crystal Set Designs by Ken Harthun
http://www.qsl.net/kc4iwt/xtal/SWMystery.htm

Short Wave Crystal Set by Al Klase
http://www.skywaves.ar88.net/SWXS/shortwave.html

Scott's Crystal Radios and links http://oldheadphones.com/crystal/index.html

Crystal Set Analysis by Phil Anderson WØXI (ARRL)
http://www.arrl.org/qex/2008/09/Anderson.pdf

High Sensitivity MOSFET Crystal Set (ARRL)
http://www.arrl.org/qst/2007/01/culter.pdf

MOSFET Crystal Radio by G1EXG http://www.creative-science.org.uk/epadmosfets.html

Peebles Crystal Radios http://www.peeblesoriginals.com/

Crystal Set Kits http://www.peeblesoriginals.com/crystal-kits.html

http://www.peeblesoriginals.com/catalog/45.php

Early Wireless http://www.earlywireless.com/crystal_sets.htm

Stay Tuned Crystal Radio Resources http://www.crystalradio.net/

Birmingham, Alabama Crystal Radio Group http://www.crystalradio.us/index.htm

Borden Radio Company - commercial website (USA)
http://www.xtalman.com/literature.html

Geoff's Crystal Radios - commercial website (UK) http://crystalreceiver.co.uk/

Crystal Radio Supply - commercialwebsite (USA)
http://www.crystalradiosupply.com/index1.html

The Xtal Set Society - commercial websites (USA) http://crystalradio.org/
http://www.midnightscience.com/

Vintage Radio http://www.vintage-radio.com/

PAGES WITH ARTICLES BY RADIO EXPERT FELIX SCERRI:

High Quality AM Receivers:

http://www.crystalradio.net/scerri/index.shtml

http://www.crystalradio.net/

http://www.xtal-sets.com/

Hi Fi AM Receiver Designs:

http://sound.westhost.com/articles/am-radio.htm

TRF Radio Websites:

Gilbert Davey's Radio Sets - A growing resource for all who either remember building radio sets to his designs or would simply like to find out more: http://www.daveysradios.org.uk

Vintage Radio - TRF Radio Designs: http://www.vintageradio.me.uk/radconnav/transtrf/

Vintage Radio - Home Page: http://www.vintageradio.me.uk/

Vintage Radio - Circuit diagrams for crystal sets, TRF and superhet radios, valve and transistor designs: http://www.vintageradio.me.uk/radconnav/radcon.htm

TRF Designs on Birmingham Alabama Crystal Radio Group: http://www.crystalradio.us/1adradios/1ad-2007-2.htm

Regen Radio Tutorial N1TEV : http://www.electronics-tutorials.com/receivers/regen-radio-receiver.htm

Coil Calculations: http://bellsouthpwp2.net/w/u/wuggy/coils.html

E.P.E. Everyday and Practical Electroncs Magazine: http://www.epemag3.com/

The Radio Board Forums - incl Audion Receiver: http://theradioboard.com/rb/viewtopic.php?t=6161&start=30

OTHER ARTICLES

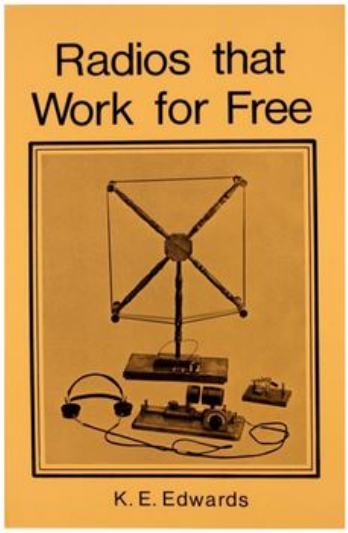
ESP by Rod Elliot - "Mad as Hell" & lead free solder directive:

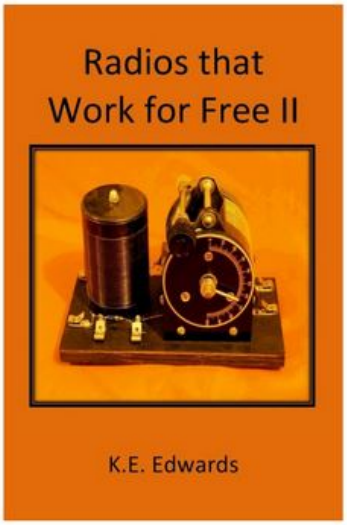
http://sound.westhost.com/madashell.htm

ESP by Rod Elliot - CFL's (Compact Fluorescent Lamps)

http://sound.westhost.com/articles/incandescent.htm

BOOKS





“Radios that Work for Free" and “Radios that Work for Free II” are now available on Amazon Kindle Books! After 38 years in print "Radios that Work for Free” is now an E Book and K. E. Edwards’ new book has just been released as an E Book as well. Crystal set builders worldwide have a new set of crystal set plans and photos to be encouraged and inspired by. These books take the reader from beginner to a competent builder of crystal sets. Take a look inside on Amazon Kindle books.
Hope And Allen Publishing - www.hopeandallen.com

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Crystal Set By Kenneth Rankin (Part 4) | **Experimental Crystal Set (Part 5)**

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What Do the Markings on Circuit Breakers Mean?

by Ed Larsen, P. E.

Have you ever been confused about what the markings on circuit breakers mean? Understanding the markings on electrical equipment is a fundamental need to ensure a safe and reliable electrical installation. Circuit breaker marking requirements are established by the requirements found in the *NEC* and the UL 489 product standard. This article will discuss the most common markings and where they can be found.

The UL 489 product standard for Molded Case Circuit Breakers specifies the information to be marked on circuit breakers and where it is to be located, so let's discuss what information needs to be marked on the circuit breaker and the location where you will find those markings. Keep in mind the UL® standard specifies minimum requirements. Circuit breaker manufacturers may provide additional information or provide information in a more convenient location.

Markings Visible without Removing Trims or Covers

UL 489 requires that some markings be visible without removing trims or covers. This location is typically referred to as the handle escutcheon (see photo 1).

Markings Visible with Trims or Covers Removed

UL 489 requires other markings be visible on an installed circuit breaker with trims or covers removed. This location is typically referred to as the face of the circuit breaker (see photos 2–4).

Other markings which should be visible with trims or covers removed are:

Independent trip – Multi-pole circuit breakers are constructed with either a common trip, where all poles are mechanically tripped when one of the poles trips, or an independent trip construction where only the pole that is involved with the overcurrent condition trips. If a 2-pole circuit breaker does not have an internal com-

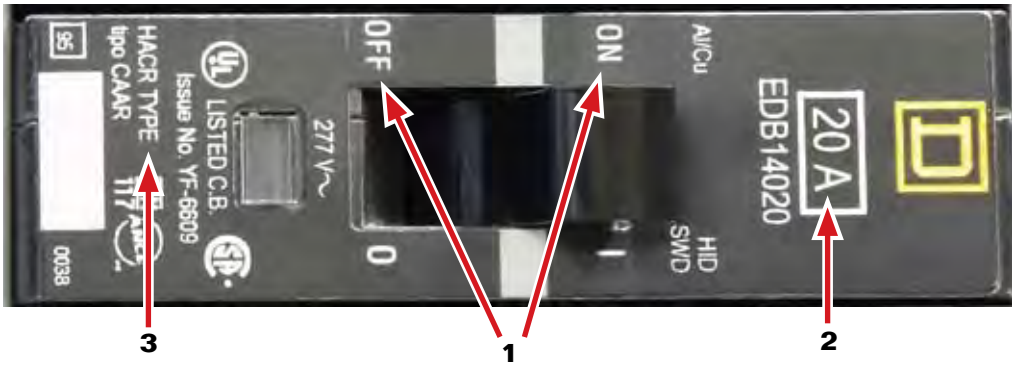

Photo 1. Markings Visible with Trims or Covers in Place

1. ON and OFF – The ON and OFF (closed and open) positions of the handle must be marked (*NEC* 240.81). These positions may also be marked with the internationally recognized "I" and "O" symbols, although this is not a UL requirement. If these markings are not visible when a motor operator is installed over the circuit breaker markings, then they must appear on the motor operator. Motor operators may be found in applications where remote or automatic operation of a circuit breaker is required.

2. Ampere rating (if 100 A or less) – The ampere rating may be located on the handle escutcheon or on the handle itself (*NEC* 240.83(B)). Circuit breakers that are rated more than 100 A may have their ampere rating marked in a position that is not visible with trims or covers in place.

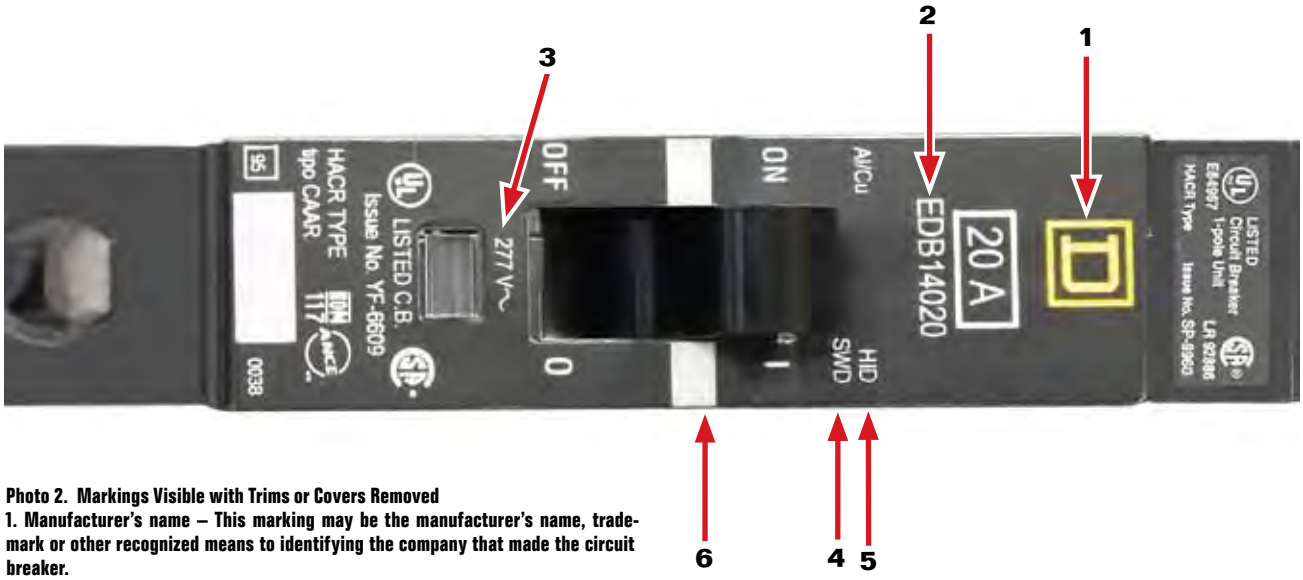

Photo 2. Markings Visible with Trims or Covers Removed

1. Manufacturer's name – This marking may be the manufacturer's name, trademark or other recognized means to identifying the company that made the circuit breaker.

2. Type designation – All circuit breakers are marked with a type designation, which may be a catalog number prefix or a separate designation. Equipment labels, such as on panelboards, will list the circuit breaker types suitable for use. Note that the word "type" may or may not be used on the circuit breaker or equipment labels. It is important to review the markings on the equipment, such as a panelboard, to make sure the circuit breaker designations on the equipment match the marking on the circuit breaker.

3. Voltage rating – All circuit breakers must be marked with a voltage rating. If the rating is not marked "ac" or "dc," then it is suitable for both. 120/240 V rated circuit breakers are suitable for use on single and three-phase 4-wire systems where the line-to-ground voltage does not exceed 120 V. Wye rated circuit breakers such as those rated 480Y/277 V, are suitable for use on three-phase 4-wire systems where the voltage to ground does not exceed 277 V. Special attention needs to be given to high leg or corner-grounded delta systems to insure that the circuit breaker has the appropriate rating. A review to see that the circuit breakers installed have a voltage rating suitable for the application is paramount for a code-compliant installation (*NEC* 240.83(E)).

4. SWD – 15- or 20-A circuit breakers rated 347 V or less may be marked "SWD," meaning that they are suitable for switching fluorescent lighting loads on a regular basis (*NEC* 240.83(D)). These circuit breakers are evaluated for high endurance use, since they will be used similar to a light switch.

5. HID – 50 A or less circuit breakers rated 480 V or less may be marked "HID," meaning they are suitable for switching high intensity discharge or fluorescent lighting loads on a regular basis. These circuit breakers may employ a different construction than a standard SWD circuit breaker in order to address the high inrush current resulting from the lower power factor created by the HID lighting (*NEC* 240.83(D)). These circuit breakers also undergo additional endurance evaluation to demonstrate their ability to perform the switching duty.

6. Trip and reset – Circuit breaker handles typically assume an intermediate position when tripped. This position must either be marked on the circuit breaker or on the equipment into which it is to be installed. If these markings are not visible when a motor operator is installed, then a "tripped" marking may appear on the motor operator.

mon trip feature, then it must be marked "Independent Trip" or "No Common Trip." *NEC* 240.20(B) is the foundational requirement for a common trip function in a circuit breaker; however, it also goes on to explain where independent trip is permitted.

For Replacement Use Only not-CTL – The Class CTL (circuit limiting) panelboard has only been in existence for about 25 years, even though the lighting and appliance branch circuit panelboard has been in the *NEC* for decades. CTL panelboards have a rejection means designed to reject more than the appropriate number of circuit breakers that can be installed in the panel. The marking "For replacement Use Only Not CTL Assemblies" means that the circuit breaker does not have CTL rejection provisions and is intended for replacement in older equipment pre-dating the CTL requirements for circuit breakers and panelboards. Circuit breakers with this designation should not be installed in a panelboard marked "Class CTL Panelboard" since that would be a violation of the listing of the assembly [*NEC* 110.3(B)].

Markings Found in Other Locations

The markings we will discuss below may appear in any location except the back of the circuit breaker. These markings include:

40°C – This marking indicates the maximum ambient temperature in which the circuit breaker can be applied at its marked ampere rating without rerating the ampacity of the circuit breaker. This marking is required for thermal-magnetic circuit breakers and is optional for electronic trip circuit breakers unless they are only suitable for a 25°C ambient, in which case they must be marked 25°C. When the ambient temperature rises above 40°C, the designer may need to consult the manufacturer to obtain rerating information (see item 4 in photo 3).

Class CTL – Circuit breakers marked Class CTL have a rejection means designed into the circuit breaker. Class CTL panelboards or assemblies, in conjunction with Class CTL circuit breakers, prevent more circuit breaker poles from being installed than the number for which the equipment is rated.

Photo 3. Markings Visible with Trims or Covers Removed

1. Line and load designation – Circuit breakers marked with "line" and "load" designations are not suitable for reverse connection. Circuit breakers with interchangeable trip units must be marked "line" and "load" unless there is no risk of shock when changing the trip unit.

2. Interrupting ratings – All circuit breakers with an interrupting rating more than 5000 A must be marked with an interrupting rating (*NEC* 240.83(C)). Interrupting ratings are stated in RMS symmetrical amperes. If the short-circuit current rating of the equipment in which the circuit breaker is installed is less than the interrupting rating of the circuit breaker, then the lesser rating applies. Circuit breakers should be reviewed after installation to ensure they have an interrupting rating suitable for the application. This marking may be found in any location except the back of circuit breakers that are 1-½ inches wide per pole or less due to the size constraint.

3. Ampere rating (if more than 100 A) – The ampere rating of a circuit breaker larger than 100 A may be found in a location that is visible after the cover or trim is removed. This marking requirement also applies to interchangeable trip units (*NEC* 240.83(A)).

5. Terminations (Cu-Al wire) – Circuit breakers must be marked with the type material (Cu-Al) and size of wire for which their terminals are suitable for use. If suitable for use with only copper or only aluminum, then the word "only" must be used. The abbreviations "CU" and "AL" are generally found on circuit breakers as permitted by the product standard. If only solid 10–14 AWG wire can be used, then that information must be noted. The number of wires per terminal will also be noted if more than one wire per terminal is permitted (*NEC* 110.14(A)).

6. Tightening torque – The nominal torque for all field-wiring terminals must be marked. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be found in any location except on the back.

7. Wire temperature ratings – Circuit breakers rated 125 A or less may be marked as suitable for use with 60°C, 60/75°C or 75°C only wire. Circuit breakers rated more than 125 A are rated for use with 75°C wire; the marking is optional. It is always permissible to use wire with a higher temperature rating, but it must be sized in accordance with the temperature marking on the circuit breaker and *NEC* Table 310.16. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be in any location except on the back (*NEC* 110.14(C)1).

8. HACR type – This marking indicates the circuit breaker is suitable for use with the group motor installations typically found in heating, air conditioning and refrigeration equipment. The *NEC* 2005 no longer has this marking requirement. The electrical industry determined that circuit breakers are considered suitable for use with such equipment without any further testing, therefore, the HACR marking is no longer required on air conditioning and refrigeration equipment or on circuit breakers for use in these applications. The requirement for this marking has also been removed from the UL 1995 product standard for HVAC equipment (see item 3 in photo 1).

Maximum wire size – Circuit breakers are typically marked with a wire range, however that marking is not mandatory. If the circuit breaker cannot accept the next larger wire size required for the ampere rating, then the maximum wire size must be marked in any location except the back (see item 5 in photo 3).

Separately shipped connectors –If connectors are not factory installed on a circuit breaker, then it must be marked with the proper connectors or terminal kits required in any location except the back (see item 8 in photo 3).

Ground-Fault Protection for People

The GFCI function, as part of a circuit breaker, provides ground-fault protection for people and has a number of unique marking and instruction requirements.

Test function – The GFCI has a test function that requires action upon installation and on a monthly basis. GFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate monthly testing.

Photo 4. 1Ø – 3Ø Marking
1. 1Ø – 3Ø marking – A 2-pole circuit breaker used to protect a 3-phase load on a corner-grounded delta system must be rated and marked for such an installation. Circuit breakers marked "1-phase – 3-phase" or "1Ø – 3Ø "are suitable for use on 3-phase corner-grounded delta or single-phase circuits (*NEC* 240.85).

Suitable for continuous operation at 100% of ampere rating only if used in a recognized enclosure 52 in. (1321 mm) by 20 in. (508 mm) x 7 1/2 in. (190 mm) deep or larger. Use only with 90° rated wire sized per the ampacity of 75° rated wire.

Figure 1. 100% Rated Marking
100 percent continuous rated – Circuit breakers are typically intended for use at not more than 80% of rated current where the load is considered continuous, or will continue for 3 hours or more (*NEC* 210.20). However, some circuit breakers are rated for continuous use at 100% of their current rating. These circuit breakers must be so marked in any location except on the back. Enclosure information such as a specific type or specific volume must also be marked. A requirement for the use of 90°C insulated wire sized to the 75°C column in *NEC* Table 310.16 and specific ventilation requirements may also be marked on the circuit breaker or equipment (*NEC* 210.20(A) and 215.3).

"Class A" marking – A "Class A" ground-fault device is intended to protect people. The Class A marking indicates that the trip threshold of the GFCI is between 4 mA and 6 mA. This marking may be in any location except the back.

Instructions – All GFCI circuit breakers must include instructions for the installer plus instructions on the use of the test function. A hangtag or self-adhesive label must also be provided, instructing the user to test the GFCI at least monthly. Inspectors should check to see that the tag or label has been properly installed.

Ground-Fault Protection for Equipment

Circuit breakers may also include a ground-fault protection for equipment (GFPE) function that, like GFCIs, has a number of unique marking and instruction requirements.

Test function – GFPE circuit breakers may have a test button or switch that may be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Trip level – GFPE circuit breakers must be marked with their trip threshold in milliamperes in a location accessible without removing trims or covers.

Instructions – All GFPE circuit breakers must include instructions for the installer.

Arc-Fault Protection

Circuit breakers may also include arc-fault protection (AFCI) that, like GFCIs, also has a number of unique marking and instruction requirements.

Device identification – AFCIs must also be identified appropriately. Branch/feeder or Combination type AFCIs must be so marked in a location visible when the trims or covers are removed. This is an important marking to note as we move into 2008, as *NEC*-2005 requires Combination AFCIs in bedrooms effective January 1, 2008 (*NEC* 210.12).

Test function – AFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Instructions – All AFCI circuit breakers must include instructions for the installer.

Circuit Breaker Markings Ensure a Safe Electrical Installation

So why are all of these markings on circuit breakers? Without them, it would be nearly impossible to install or inspect an installation for the appropriate performance ratings and fundamental electrical connections. When designing or completing an installation, key items to review are:

1. Are the voltage, continuous current, and interrupting ratings appropriate for the application?
2. Does the application require SWD or HID ratings?
3. Is the wire type and size appropriate for the circuit breaker?
4. Is the circuit breaker suitable for the equipment in which it is installed? Have other protective functions such as GFCI or AFCI been provided as required by the *NEC*?
5. Is the temperature rating of the circuit breaker suitable for the application?

The UL Marking Guide for Molded Case Circuit Breakers is a valuable resource to understand circuit breaker markings that may further explain these and other markings in detail. If you have questions about CB markings not answered here, consult the Marking Guide or the manufacturer to assist in an *NEC*-compliant installation.

Mr. Larsen, P. E., is the manager, Industry Standards for Circuit Protection for Square D Company / Schneider Electric. Mr. Larsen has served in various positions over the past 32 years. He is a member of several UL, IEEE and NEMA technical committees. His other memberships include NFPA, IAEI and CANENA.

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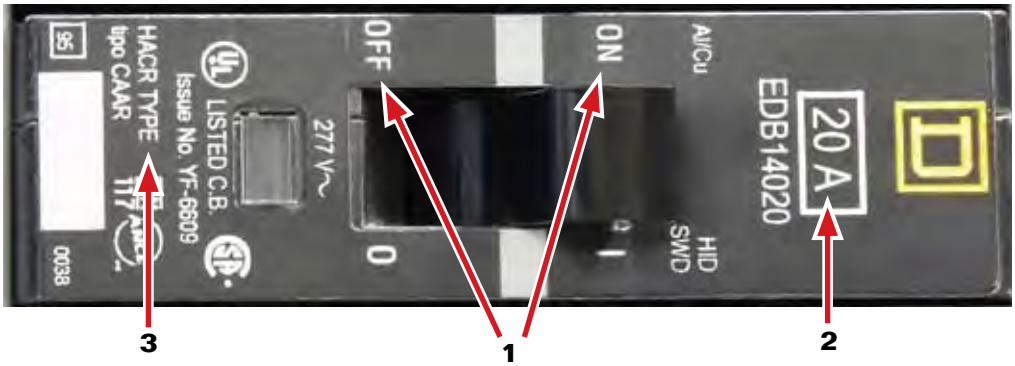

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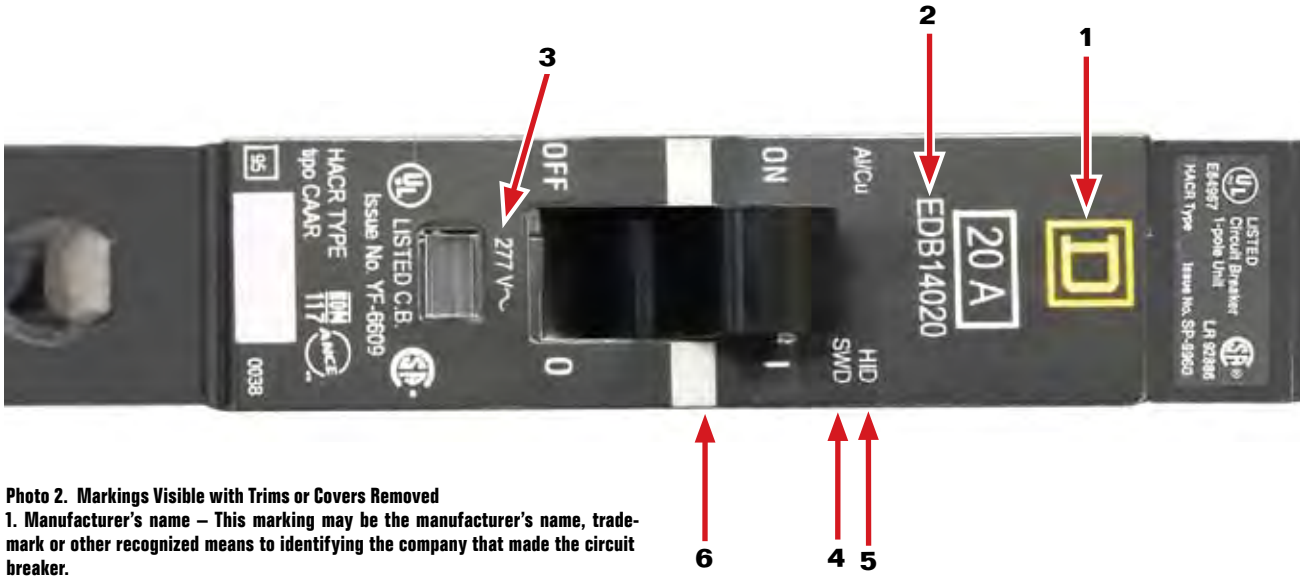


Photo 2. Markings Visible with Trims or Covers Removed

1. Manufacturer's name – This marking may be the manufacturer's name, trademark or other recognized means to identifying the company that made the circuit breaker.

2. Type designation – All circuit breakers are marked with a type designation, which may be a catalog number prefix or a separate designation. Equipment labels, such as on panelboards, will list the circuit breaker types suitable for use. Note that the word "type" may or may not be used on the circuit breaker or equipment labels. It is important to review the markings on the equipment, such as a panelboard, to make sure the circuit breaker designations on the equipment match the marking on the circuit breaker.

3. Voltage rating – All circuit breakers must be marked with a voltage rating. If the rating is not marked "ac" or "dc," then it is suitable for both. 120/240 V rated circuit breakers are suitable for use on single and three-phase 4-wire systems where the line-to-ground voltage does not exceed 120 V. Wye rated circuit breakers such as those rated 480Y/277 V, are suitable for use on three-phase 4-wire systems where the voltage to ground does not exceed 277 V. Special attention needs to be given to high leg or corner-grounded delta systems to insure that the circuit breaker has the appropriate rating. A review to see that the circuit breakers installed have a voltage rating suitable for the application is paramount for a code-compliant installation (*NEC* 240.83(E)).

4. SWD – 15- or 20-A circuit breakers rated 347 V or less may be marked "SWD," meaning that they are suitable for switching fluorescent lighting loads on a regular basis (*NEC* 240.83(D)). These circuit breakers are evaluated for high endurance use, since they will be used similar to a light switch.

5. HID – 50 A or less circuit breakers rated 480 V or less may be marked "HID," meaning they are suitable for switching high intensity discharge or fluorescent lighting loads on a regular basis. These circuit breakers may employ a different construction than a standard SWD circuit breaker in order to address the high inrush current resulting from the lower power factor created by the HID lighting (*NEC* 240.83(D)). These circuit breakers also undergo additional endurance evaluation to demonstrate their ability to perform the switching duty.

6. Trip and reset – Circuit breaker handles typically assume an intermediate position when tripped. This position must either be marked on the circuit breaker or on the equipment into which it is to be installed. If these markings are not visible when a motor operator is installed, then a "tripped" marking may appear on the motor operator.

mon trip feature, then it must be marked "Independent Trip" or "No Common Trip." *NEC* 240.20(B) is the foundational requirement for a common trip function in a circuit breaker; however, it also goes on to explain where independent trip is permitted.

For Replacement Use Only not-CTL – The Class CTL (circuit limiting) panelboard has only been in existence for about 25 years, even though the lighting and appliance branch circuit panelboard has been in the *NEC* for decades. CTL panelboards have a rejection means designed to reject more than the appropriate number of circuit breakers that can be installed in the panel. The marking "For replacement Use Only Not CTL Assemblies" means that the circuit breaker does not have CTL rejection provisions and is intended for replacement in older equipment pre-dating the CTL requirements for circuit breakers and panelboards. Circuit breakers with this designation should not be installed in a panelboard marked "Class CTL Panelboard" since that would be a violation of the listing of the assembly [*NEC* 110.3(B)].

Markings Found in Other Locations

The markings we will discuss below may appear in any location except the back of the circuit breaker. These markings include:

40°C – This marking indicates the maximum ambient temperature in which the circuit breaker can be applied at its marked ampere rating without rerating the ampacity of the circuit breaker. This marking is required for thermal-magnetic circuit breakers and is optional for electronic trip circuit breakers unless they are only suitable for a 25°C ambient, in which case they must be marked 25°C. When the ambient temperature rises above 40°C, the designer may need to consult the manufacturer to obtain rerating information (see item 4 in photo 3).

Class CTL – Circuit breakers marked Class CTL have a rejection means designed into the circuit breaker. Class CTL panelboards or assemblies, in conjunction with Class CTL circuit breakers, prevent more circuit breaker poles from being installed than the number for which the equipment is rated.

Photo 3. Markings Visible with Trims or Covers Removed

1. Line and load designation – Circuit breakers marked with "line" and "load" designations are not suitable for reverse connection. Circuit breakers with interchangeable trip units must be marked "line" and "load" unless there is no risk of shock when changing the trip unit.

2. Interrupting ratings – All circuit breakers with an interrupting rating more than 5000 A must be marked with an interrupting rating (*NEC* 240.83(C)). Interrupting ratings are stated in RMS symmetrical amperes. If the short-circuit current rating of the equipment in which the circuit breaker is installed is less than the interrupting rating of the circuit breaker, then the lesser rating applies. Circuit breakers should be reviewed after installation to ensure they have an interrupting rating suitable for the application. This marking may be found in any location except the back of circuit breakers that are 1-½ inches wide per pole or less due to the size constraint.

3. Ampere rating (if more than 100 A) – The ampere rating of a circuit breaker larger than 100 A may be found in a location that is visible after the cover or trim is removed. This marking requirement also applies to interchangeable trip units (*NEC* 240.83(A)).

5. Terminations (Cu-Al wire) – Circuit breakers must be marked with the type material (Cu-Al) and size of wire for which their terminals are suitable for use. If suitable for use with only copper or only aluminum, then the word "only" must be used. The abbreviations "CU" and "AL" are generally found on circuit breakers as permitted by the product standard. If only solid 10–14 AWG wire can be used, then that information must be noted. The number of wires per terminal will also be noted if more than one wire per terminal is permitted (*NEC* 110.14(A)).

6. Tightening torque – The nominal torque for all field-wiring terminals must be marked. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be found in any location except on the back.

7. Wire temperature ratings – Circuit breakers rated 125 A or less may be marked as suitable for use with 60°C, 60/75°C or 75°C only wire. Circuit breakers rated more than 125 A are rated for use with 75°C wire; the marking is optional. It is always permissible to use wire with a higher temperature rating, but it must be sized in accordance with the temperature marking on the circuit breaker and *NEC* Table 310.16. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be in any location except on the back (*NEC* 110.14(C)1).

8. HACR type – This marking indicates the circuit breaker is suitable for use with the group motor installations typically found in heating, air conditioning and refrigeration equipment. The *NEC* 2005 no longer has this marking requirement. The electrical industry determined that circuit breakers are considered suitable for use with such equipment without any further testing, therefore, the HACR marking is no longer required on air conditioning and refrigeration equipment or on circuit breakers for use in these applications. The requirement for this marking has also been removed from the UL 1995 product standard for HVAC equipment (see item 3 in photo 1).

Maximum wire size – Circuit breakers are typically marked with a wire range, however that marking is not mandatory. If the circuit breaker cannot accept the next larger wire size required for the ampere rating, then the maximum wire size must be marked in any location except the back (see item 5 in photo 3).

Separately shipped connectors –If connectors are not factory installed on a circuit breaker, then it must be marked with the proper connectors or terminal kits required in any location except the back (see item 8 in photo 3).

Ground-Fault Protection for People

The GFCI function, as part of a circuit breaker, provides ground-fault protection for people and has a number of unique marking and instruction requirements.

Test function – The GFCI has a test function that requires action upon installation and on a monthly basis. GFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate monthly testing.

Photo 4. 1Ø – 3Ø Marking
1. 1Ø – 3Ø marking – A 2-pole circuit breaker used to protect a 3-phase load on a corner-grounded delta system must be rated and marked for such an installation. Circuit breakers marked "1-phase – 3-phase" or "1Ø – 3Ø "are suitable for use on 3-phase corner-grounded delta or single-phase circuits (*NEC* 240.85).

Suitable for continuous operation at 100% of ampere rating only if used in a recognized enclosure 52 in. (1321 mm) by 20 in. (508 mm) x 7 1/2 in. (190 mm) deep or larger. Use only with 90° rated wire sized per the ampacity of 75° rated wire.

Figure 1. 100% Rated Marking
100 percent continuous rated – Circuit breakers are typically intended for use at not more than 80% of rated current where the load is considered continuous, or will continue for 3 hours or more (*NEC* 210.20). However, some circuit breakers are rated for continuous use at 100% of their current rating. These circuit breakers must be so marked in any location except on the back. Enclosure information such as a specific type or specific volume must also be marked. A requirement for the use of 90°C insulated wire sized to the 75°C column in *NEC* Table 310.16 and specific ventilation requirements may also be marked on the circuit breaker or equipment (*NEC* 210.20(A) and 215.3).

"Class A" marking – A "Class A" ground-fault device is intended to protect people. The Class A marking indicates that the trip threshold of the GFCI is between 4 mA and 6 mA. This marking may be in any location except the back.

Instructions – All GFCI circuit breakers must include instructions for the installer plus instructions on the use of the test function. A hangtag or self-adhesive label must also be provided, instructing the user to test the GFCI at least monthly. Inspectors should check to see that the tag or label has been properly installed.

Ground-Fault Protection for Equipment

Circuit breakers may also include a ground-fault protection for equipment (GFPE) function that, like GFCIs, has a number of unique marking and instruction requirements.

Test function – GFPE circuit breakers may have a test button or switch that may be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Trip level – GFPE circuit breakers must be marked with their trip threshold in milliamperes in a location accessible without removing trims or covers.

Instructions – All GFPE circuit breakers must include instructions for the installer.

Arc-Fault Protection

Circuit breakers may also include arc-fault protection (AFCI) that, like GFCIs, also has a number of unique marking and instruction requirements.

Device identification – AFCIs must also be identified appropriately. Branch/feeder or Combination type AFCIs must be so marked in a location visible when the trims or covers are removed. This is an important marking to note as we move into 2008, as *NEC*-2005 requires Combination AFCIs in bedrooms effective January 1, 2008 (*NEC* 210.12).

Test function – AFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Instructions – All AFCI circuit breakers must include instructions for the installer.

Circuit Breaker Markings Ensure a Safe Electrical Installation

So why are all of these markings on circuit breakers? Without them, it would be nearly impossible to install or inspect an installation for the appropriate performance ratings and fundamental electrical connections. When designing or completing an installation, key items to review are:

1. Are the voltage, continuous current, and interrupting ratings appropriate for the application?
2. Does the application require SWD or HID ratings?
3. Is the wire type and size appropriate for the circuit breaker?
4. Is the circuit breaker suitable for the equipment in which it is installed? Have other protective functions such as GFCI or AFCI been provided as required by the *NEC*?
5. Is the temperature rating of the circuit breaker suitable for the application?

The UL Marking Guide for Molded Case Circuit Breakers is a valuable resource to understand circuit breaker markings that may further explain these and other markings in detail. If you have questions about CB markings not answered here, consult the Marking Guide or the manufacturer to assist in an *NEC*-compliant installation.

Mr. Larsen, P. E., is the manager, Industry Standards for Circuit Protection for Square D Company / Schneider Electric. Mr. Larsen has served in various positions over the past 32 years. He is a member of several UL, IEEE and NEMA technical committees. His other memberships include NFPA, IAEI and CANENA.

What Do the Markings on Circuit Breakers Mean?

by Ed Larsen, P. E.

Have you ever been confused about what the markings on circuit breakers mean? Understanding the markings on electrical equipment is a fundamental need to ensure a safe and reliable electrical installation. Circuit breaker marking requirements are established by the requirements found in the *NEC* and the UL 489 product standard. This article will discuss the most common markings and where they can be found.

The UL 489 product standard for Molded Case Circuit Breakers specifies the information to be marked on circuit breakers and where it is to be located, so let's discuss what information needs to be marked on the circuit breaker and the location where you will find those markings. Keep in mind the UL® standard specifies minimum requirements. Circuit breaker manufacturers may provide additional information or provide information in a more convenient location.

Markings Visible without Removing Trims or Covers

UL 489 requires that some markings be visible without removing trims or covers. This location is typically referred to as the handle escutcheon (see photo 1).

Markings Visible with Trims or Covers Removed

UL 489 requires other markings be visible on an installed circuit breaker with trims or covers removed. This location is typically referred to as the face of the circuit breaker (see photos 2–4).

Other markings which should be visible with trims or covers removed are:

Independent trip – Multi-pole circuit breakers are constructed with either a common trip, where all poles are mechanically tripped when one of the poles trips, or an independent trip construction where only the pole that is involved with the overcurrent condition trips. If a 2-pole circuit breaker does not have an internal com-

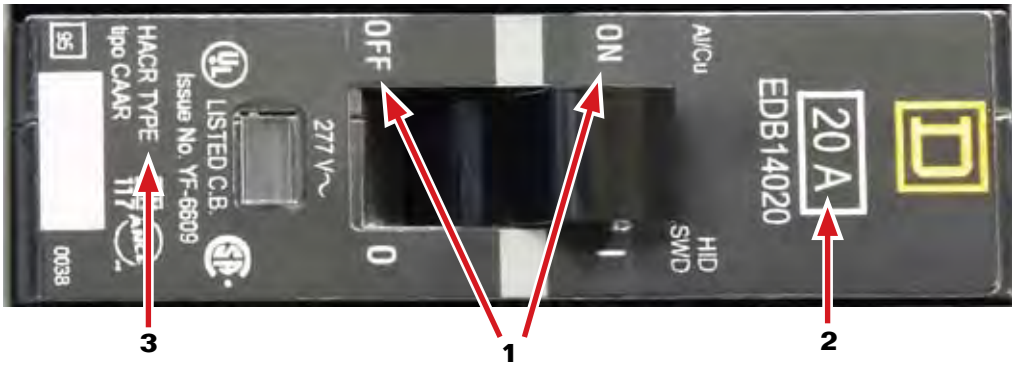
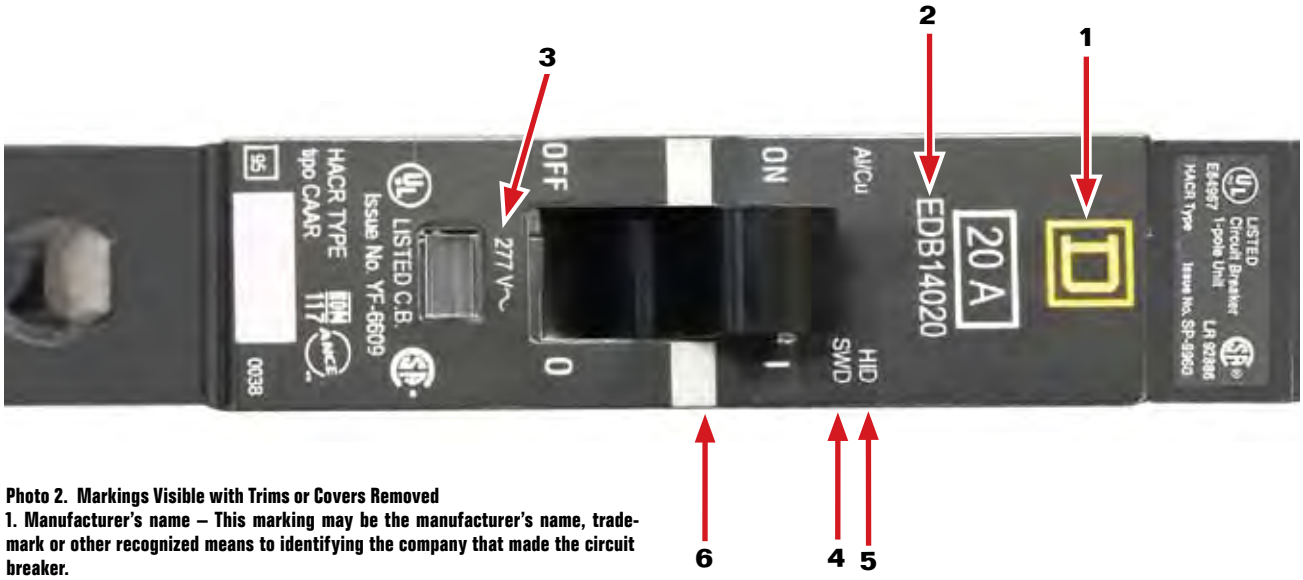


Photo 1. Markings Visible with Trims or Covers in Place

1. ON and OFF – The ON and OFF (closed and open) positions of the handle must be marked (*NEC* 240.81). These positions may also be marked with the internationally recognized "I" and "O" symbols, although this is not a UL requirement. If these markings are not visible when a motor operator is installed over the circuit breaker markings, then they must appear on the motor operator. Motor operators may be found in applications where remote or automatic operation of a circuit breaker is required.

2. Ampere rating (if 100 A or less) – The ampere rating may be located on the handle escutcheon or on the handle itself (*NEC* 240.83(B)). Circuit breakers that are rated more than 100 A may have their ampere rating marked in a position that is not visible with trims or covers in place.



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1. Manufacturer's name – This marking may be the manufacturer's name, trademark or other recognized means to identifying the company that made the circuit breaker.

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3. Voltage rating – All circuit breakers must be marked with a voltage rating. If the rating is not marked "ac" or "dc," then it is suitable for both. 120/240 V rated circuit breakers are suitable for use on single and three-phase 4-wire systems where the line-to-ground voltage does not exceed 120 V. Wye rated circuit breakers such as those rated 480Y/277 V, are suitable for use on three-phase 4-wire systems where the voltage to ground does not exceed 277 V. Special attention needs to be given to high leg or corner-grounded delta systems to insure that the circuit breaker has the appropriate rating. A review to see that the circuit breakers installed have a voltage rating suitable for the application is paramount for a code-compliant installation (*NEC* 240.83(E)).

4. SWD – 15- or 20-A circuit breakers rated 347 V or less may be marked "SWD," meaning that they are suitable for switching fluorescent lighting loads on a regular basis (*NEC* 240.83(D)). These circuit breakers are evaluated for high endurance use, since they will be used similar to a light switch.

5. HID – 50 A or less circuit breakers rated 480 V or less may be marked "HID," meaning they are suitable for switching high intensity discharge or fluorescent lighting loads on a regular basis. These circuit breakers may employ a different construction than a standard SWD circuit breaker in order to address the high inrush current resulting from the lower power factor created by the HID lighting (*NEC* 240.83(D)). These circuit breakers also undergo additional endurance evaluation to demonstrate their ability to perform the switching duty.

6. Trip and reset – Circuit breaker handles typically assume an intermediate position when tripped. This position must either be marked on the circuit breaker or on the equipment into which it is to be installed. If these markings are not visible when a motor operator is installed, then a "tripped" marking may appear on the motor operator.

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Markings Found in Other Locations

The markings we will discuss below may appear in any location except the back of the circuit breaker. These markings include:

40°C – This marking indicates the maximum ambient temperature in which the circuit breaker can be applied at its marked ampere rating without rerating the ampacity of the circuit breaker. This marking is required for thermal-magnetic circuit breakers and is optional for electronic trip circuit breakers unless they are only suitable for a 25°C ambient, in which case they must be marked 25°C. When the ambient temperature rises above 40°C, the designer may need to consult the manufacturer to obtain rerating information (see item 4 in photo 3).

Class CTL – Circuit breakers marked Class CTL have a rejection means designed into the circuit breaker. Class CTL panelboards or assemblies, in conjunction with Class CTL circuit breakers, prevent more circuit breaker poles from being installed than the number for which the equipment is rated.

Photo 3. Markings Visible with Trims or Covers Removed

1. Line and load designation – Circuit breakers marked with "line" and "load" designations are not suitable for reverse connection. Circuit breakers with interchangeable trip units must be marked "line" and "load" unless there is no risk of shock when changing the trip unit.

2. Interrupting ratings – All circuit breakers with an interrupting rating more than 5000 A must be marked with an interrupting rating (*NEC* 240.83(C)). Interrupting ratings are stated in RMS symmetrical amperes. If the short-circuit current rating of the equipment in which the circuit breaker is installed is less than the interrupting rating of the circuit breaker, then the lesser rating applies. Circuit breakers should be reviewed after installation to ensure they have an interrupting rating suitable for the application. This marking may be found in any location except the back of circuit breakers that are 1-½ inches wide per pole or less due to the size constraint.

3. Ampere rating (if more than 100 A) – The ampere rating of a circuit breaker larger than 100 A may be found in a location that is visible after the cover or trim is removed. This marking requirement also applies to interchangeable trip units (*NEC* 240.83(A)).

5. Terminations (Cu-Al wire) – Circuit breakers must be marked with the type material (Cu-Al) and size of wire for which their terminals are suitable for use. If suitable for use with only copper or only aluminum, then the word "only" must be used. The abbreviations "CU" and "AL" are generally found on circuit breakers as permitted by the product standard. If only solid 10–14 AWG wire can be used, then that information must be noted. The number of wires per terminal will also be noted if more than one wire per terminal is permitted (*NEC* 110.14(A)).

6. Tightening torque – The nominal torque for all field-wiring terminals must be marked. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be found in any location except on the back.

7. Wire temperature ratings – Circuit breakers rated 125 A or less may be marked as suitable for use with 60°C, 60/75°C or 75°C only wire. Circuit breakers rated more than 125 A are rated for use with 75°C wire; the marking is optional. It is always permissible to use wire with a higher temperature rating, but it must be sized in accordance with the temperature marking on the circuit breaker and *NEC* Table 310.16. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be in any location except on the back (*NEC* 110.14(C)1).

8. HACR type – This marking indicates the circuit breaker is suitable for use with the group motor installations typically found in heating, air conditioning and refrigeration equipment. The *NEC* 2005 no longer has this marking requirement. The electrical industry determined that circuit breakers are considered suitable for use with such equipment without any further testing, therefore, the HACR marking is no longer required on air conditioning and refrigeration equipment or on circuit breakers for use in these applications. The requirement for this marking has also been removed from the UL 1995 product standard for HVAC equipment (see item 3 in photo 1).

Maximum wire size – Circuit breakers are typically marked with a wire range, however that marking is not mandatory. If the circuit breaker cannot accept the next larger wire size required for the ampere rating, then the maximum wire size must be marked in any location except the back (see item 5 in photo 3).

Separately shipped connectors –If connectors are not factory installed on a circuit breaker, then it must be marked with the proper connectors or terminal kits required in any location except the back (see item 8 in photo 3).

Ground-Fault Protection for People

The GFCI function, as part of a circuit breaker, provides ground-fault protection for people and has a number of unique marking and instruction requirements.

Test function – The GFCI has a test function that requires action upon installation and on a monthly basis. GFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate monthly testing.

Photo 4. 1Ø – 3Ø Marking
1. 1Ø – 3Ø marking – A 2-pole circuit breaker used to protect a 3-phase load on a corner-grounded delta system must be rated and marked for such an installation. Circuit breakers marked "1-phase – 3-phase" or "1Ø – 3Ø "are suitable for use on 3-phase corner-grounded delta or single-phase circuits (*NEC* 240.85).

Suitable for continuous operation at 100% of ampere rating only if used in a recognized enclosure 52 in. (1321 mm) by 20 in. (508 mm) x 7 1/2 in. (190 mm) deep or larger. Use only with 90° rated wire sized per the ampacity of 75° rated wire.

Figure 1. 100% Rated Marking
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Instructions – All GFCI circuit breakers must include instructions for the installer plus instructions on the use of the test function. A hangtag or self-adhesive label must also be provided, instructing the user to test the GFCI at least monthly. Inspectors should check to see that the tag or label has been properly installed.

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Trip level – GFPE circuit breakers must be marked with their trip threshold in milliamperes in a location accessible without removing trims or covers.

Instructions – All GFPE circuit breakers must include instructions for the installer.

Arc-Fault Protection

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Device identification – AFCIs must also be identified appropriately. Branch/feeder or Combination type AFCIs must be so marked in a location visible when the trims or covers are removed. This is an important marking to note as we move into 2008, as *NEC*-2005 requires Combination AFCIs in bedrooms effective January 1, 2008 (*NEC* 210.12).

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UL 489 requires that some markings be visible without removing trims or covers. This location is typically referred to as the handle escutcheon (see photo 1).

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UL 489 requires other markings be visible on an installed circuit breaker with trims or covers removed. This location is typically referred to as the face of the circuit breaker (see photos 2–4).

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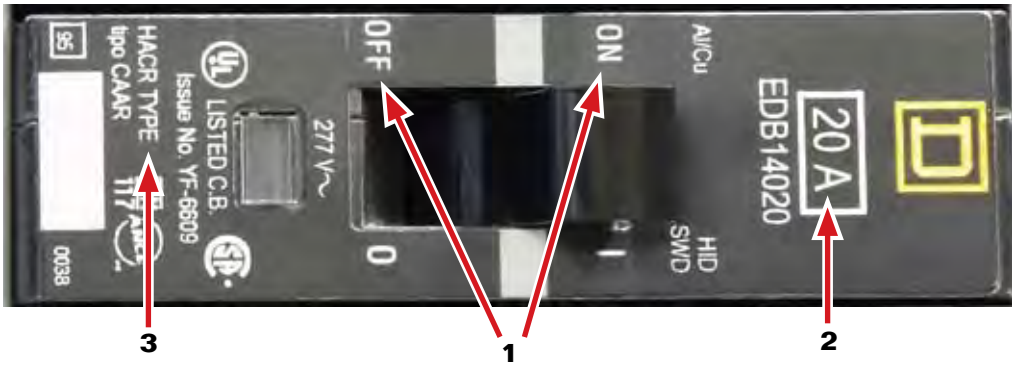


Photo 1. Markings Visible with Trims or Covers in Place

1. ON and OFF – The ON and OFF (closed and open) positions of the handle must be marked (*NEC* 240.81). These positions may also be marked with the internationally recognized "I" and "O" symbols, although this is not a UL requirement. If these markings are not visible when a motor operator is installed over the circuit breaker markings, then they must appear on the motor operator. Motor operators may be found in applications where remote or automatic operation of a circuit breaker is required.

2. Ampere rating (if 100 A or less) – The ampere rating may be located on the handle escutcheon or on the handle itself (*NEC* 240.83(B)). Circuit breakers that are rated more than 100 A may have their ampere rating marked in a position that is not visible with trims or covers in place.

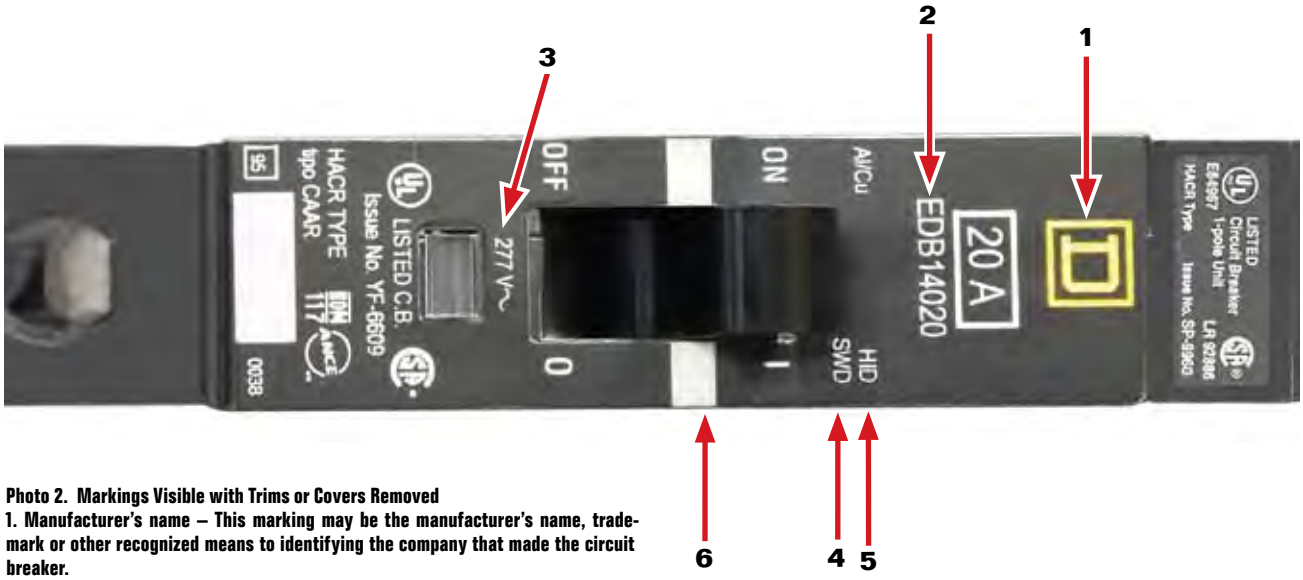


Photo 2. Markings Visible with Trims or Covers Removed

1. Manufacturer's name – This marking may be the manufacturer's name, trademark or other recognized means to identifying the company that made the circuit breaker.

2. Type designation – All circuit breakers are marked with a type designation, which may be a catalog number prefix or a separate designation. Equipment labels, such as on panelboards, will list the circuit breaker types suitable for use. Note that the word "type" may or may not be used on the circuit breaker or equipment labels. It is important to review the markings on the equipment, such as a panelboard, to make sure the circuit breaker designations on the equipment match the marking on the circuit breaker.

3. Voltage rating – All circuit breakers must be marked with a voltage rating. If the rating is not marked "ac" or "dc," then it is suitable for both. 120/240 V rated circuit breakers are suitable for use on single and three-phase 4-wire systems where the line-to-ground voltage does not exceed 120 V. Wye rated circuit breakers such as those rated 480Y/277 V, are suitable for use on three-phase 4-wire systems where the voltage to ground does not exceed 277 V. Special attention needs to be given to high leg or corner-grounded delta systems to insure that the circuit breaker has the appropriate rating. A review to see that the circuit breakers installed have a voltage rating suitable for the application is paramount for a code-compliant installation (*NEC* 240.83(E)).

4. SWD – 15- or 20-A circuit breakers rated 347 V or less may be marked "SWD," meaning that they are suitable for switching fluorescent lighting loads on a regular basis (*NEC* 240.83(D)). These circuit breakers are evaluated for high endurance use, since they will be used similar to a light switch.

5. HID – 50 A or less circuit breakers rated 480 V or less may be marked "HID," meaning they are suitable for switching high intensity discharge or fluorescent lighting loads on a regular basis. These circuit breakers may employ a different construction than a standard SWD circuit breaker in order to address the high inrush current resulting from the lower power factor created by the HID lighting (*NEC* 240.83(D)). These circuit breakers also undergo additional endurance evaluation to demonstrate their ability to perform the switching duty.

6. Trip and reset – Circuit breaker handles typically assume an intermediate position when tripped. This position must either be marked on the circuit breaker or on the equipment into which it is to be installed. If these markings are not visible when a motor operator is installed, then a "tripped" marking may appear on the motor operator.

mon trip feature, then it must be marked "Independent Trip" or "No Common Trip." *NEC* 240.20(B) is the foundational requirement for a common trip function in a circuit breaker; however, it also goes on to explain where independent trip is permitted.

For Replacement Use Only not-CTL – The Class CTL (circuit limiting) panelboard has only been in existence for about 25 years, even though the lighting and appliance branch circuit panelboard has been in the *NEC* for decades. CTL panelboards have a rejection means designed to reject more than the appropriate number of circuit breakers that can be installed in the panel. The marking "For replacement Use Only Not CTL Assemblies" means that the circuit breaker does not have CTL rejection provisions and is intended for replacement in older equipment pre-dating the CTL requirements for circuit breakers and panelboards. Circuit breakers with this designation should not be installed in a panelboard marked "Class CTL Panelboard" since that would be a violation of the listing of the assembly [*NEC* 110.3(B)].

Markings Found in Other Locations

The markings we will discuss below may appear in any location except the back of the circuit breaker. These markings include:

40°C – This marking indicates the maximum ambient temperature in which the circuit breaker can be applied at its marked ampere rating without rerating the ampacity of the circuit breaker. This marking is required for thermal-magnetic circuit breakers and is optional for electronic trip circuit breakers unless they are only suitable for a 25°C ambient, in which case they must be marked 25°C. When the ambient temperature rises above 40°C, the designer may need to consult the manufacturer to obtain rerating information (see item 4 in photo 3).

Class CTL – Circuit breakers marked Class CTL have a rejection means designed into the circuit breaker. Class CTL panelboards or assemblies, in conjunction with Class CTL circuit breakers, prevent more circuit breaker poles from being installed than the number for which the equipment is rated.

Photo 3. Markings Visible with Trims or Covers Removed

1. Line and load designation – Circuit breakers marked with "line" and "load" designations are not suitable for reverse connection. Circuit breakers with interchangeable trip units must be marked "line" and "load" unless there is no risk of shock when changing the trip unit.

2. Interrupting ratings – All circuit breakers with an interrupting rating more than 5000 A must be marked with an interrupting rating (*NEC* 240.83(C)). Interrupting ratings are stated in RMS symmetrical amperes. If the short-circuit current rating of the equipment in which the circuit breaker is installed is less than the interrupting rating of the circuit breaker, then the lesser rating applies. Circuit breakers should be reviewed after installation to ensure they have an interrupting rating suitable for the application. This marking may be found in any location except the back of circuit breakers that are 1-½ inches wide per pole or less due to the size constraint.

3. Ampere rating (if more than 100 A) – The ampere rating of a circuit breaker larger than 100 A may be found in a location that is visible after the cover or trim is removed. This marking requirement also applies to interchangeable trip units (*NEC* 240.83(A)).

5. Terminations (Cu-Al wire) – Circuit breakers must be marked with the type material (Cu-Al) and size of wire for which their terminals are suitable for use. If suitable for use with only copper or only aluminum, then the word "only" must be used. The abbreviations "CU" and "AL" are generally found on circuit breakers as permitted by the product standard. If only solid 10–14 AWG wire can be used, then that information must be noted. The number of wires per terminal will also be noted if more than one wire per terminal is permitted (*NEC* 110.14(A)).

6. Tightening torque – The nominal torque for all field-wiring terminals must be marked. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be found in any location except on the back.

7. Wire temperature ratings – Circuit breakers rated 125 A or less may be marked as suitable for use with 60°C, 60/75°C or 75°C only wire. Circuit breakers rated more than 125 A are rated for use with 75°C wire; the marking is optional. It is always permissible to use wire with a higher temperature rating, but it must be sized in accordance with the temperature marking on the circuit breaker and *NEC* Table 310.16. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be in any location except on the back (*NEC* 110.14(C)1).

8. HACR type – This marking indicates the circuit breaker is suitable for use with the group motor installations typically found in heating, air conditioning and refrigeration equipment. The *NEC* 2005 no longer has this marking requirement. The electrical industry determined that circuit breakers are considered suitable for use with such equipment without any further testing, therefore, the HACR marking is no longer required on air conditioning and refrigeration equipment or on circuit breakers for use in these applications. The requirement for this marking has also been removed from the UL 1995 product standard for HVAC equipment (see item 3 in photo 1).

Maximum wire size – Circuit breakers are typically marked with a wire range, however that marking is not mandatory. If the circuit breaker cannot accept the next larger wire size required for the ampere rating, then the maximum wire size must be marked in any location except the back (see item 5 in photo 3).

Separately shipped connectors –If connectors are not factory installed on a circuit breaker, then it must be marked with the proper connectors or terminal kits required in any location except the back (see item 8 in photo 3).

Ground-Fault Protection for People

The GFCI function, as part of a circuit breaker, provides ground-fault protection for people and has a number of unique marking and instruction requirements.

Test function – The GFCI has a test function that requires action upon installation and on a monthly basis. GFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate monthly testing.

Photo 4. 1Ø – 3Ø Marking
1. 1Ø – 3Ø marking – A 2-pole circuit breaker used to protect a 3-phase load on a corner-grounded delta system must be rated and marked for such an installation. Circuit breakers marked "1-phase – 3-phase" or "1Ø – 3Ø "are suitable for use on 3-phase corner-grounded delta or single-phase circuits (*NEC* 240.85).

Suitable for continuous operation at 100% of ampere rating only if used in a recognized enclosure 52 in. (1321 mm) by 20 in. (508 mm) x 7 1/2 in. (190 mm) deep or larger. Use only with 90° rated wire sized per the ampacity of 75° rated wire.

Figure 1. 100% Rated Marking
100 percent continuous rated – Circuit breakers are typically intended for use at not more than 80% of rated current where the load is considered continuous, or will continue for 3 hours or more (*NEC* 210.20). However, some circuit breakers are rated for continuous use at 100% of their current rating. These circuit breakers must be so marked in any location except on the back. Enclosure information such as a specific type or specific volume must also be marked. A requirement for the use of 90°C insulated wire sized to the 75°C column in *NEC* Table 310.16 and specific ventilation requirements may also be marked on the circuit breaker or equipment (*NEC* 210.20(A) and 215.3).

"Class A" marking – A "Class A" ground-fault device is intended to protect people. The Class A marking indicates that the trip threshold of the GFCI is between 4 mA and 6 mA. This marking may be in any location except the back.

Instructions – All GFCI circuit breakers must include instructions for the installer plus instructions on the use of the test function. A hangtag or self-adhesive label must also be provided, instructing the user to test the GFCI at least monthly. Inspectors should check to see that the tag or label has been properly installed.

Ground-Fault Protection for Equipment

Circuit breakers may also include a ground-fault protection for equipment (GFPE) function that, like GFCIs, has a number of unique marking and instruction requirements.

Test function – GFPE circuit breakers may have a test button or switch that may be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Trip level – GFPE circuit breakers must be marked with their trip threshold in milliamperes in a location accessible without removing trims or covers.

Instructions – All GFPE circuit breakers must include instructions for the installer.

Arc-Fault Protection

Circuit breakers may also include arc-fault protection (AFCI) that, like GFCIs, also has a number of unique marking and instruction requirements.

Device identification – AFCIs must also be identified appropriately. Branch/feeder or Combination type AFCIs must be so marked in a location visible when the trims or covers are removed. This is an important marking to note as we move into 2008, as *NEC*-2005 requires Combination AFCIs in bedrooms effective January 1, 2008 (*NEC* 210.12).

Test function – AFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Instructions – All AFCI circuit breakers must include instructions for the installer.

Circuit Breaker Markings Ensure a Safe Electrical Installation

So why are all of these markings on circuit breakers? Without them, it would be nearly impossible to install or inspect an installation for the appropriate performance ratings and fundamental electrical connections. When designing or completing an installation, key items to review are:

1. Are the voltage, continuous current, and interrupting ratings appropriate for the application?
2. Does the application require SWD or HID ratings?
3. Is the wire type and size appropriate for the circuit breaker?
4. Is the circuit breaker suitable for the equipment in which it is installed? Have other protective functions such as GFCI or AFCI been provided as required by the *NEC*?
5. Is the temperature rating of the circuit breaker suitable for the application?

The UL Marking Guide for Molded Case Circuit Breakers is a valuable resource to understand circuit breaker markings that may further explain these and other markings in detail. If you have questions about CB markings not answered here, consult the Marking Guide or the manufacturer to assist in an *NEC*-compliant installation.

Mr. Larsen, P. E., is the manager, Industry Standards for Circuit Protection for Square D Company / Schneider Electric. Mr. Larsen has served in various positions over the past 32 years. He is a member of several UL, IEEE and NEMA technical committees. His other memberships include NFPA, IAEI and CANENA.

What Do the Markings on Circuit Breakers Mean?

by Ed Larsen, P. E.

Have you ever been confused about what the markings on circuit breakers mean? Understanding the markings on electrical equipment is a fundamental need to ensure a safe and reliable electrical installation. Circuit breaker marking requirements are established by the requirements found in the *NEC* and the UL 489 product standard. This article will discuss the most common markings and where they can be found.

The UL 489 product standard for Molded Case Circuit Breakers specifies the information to be marked on circuit breakers and where it is to be located, so let's discuss what information needs to be marked on the circuit breaker and the location where you will find those markings. Keep in mind the UL® standard specifies minimum requirements. Circuit breaker manufacturers may provide additional information or provide information in a more convenient location.

Markings Visible without Removing Trims or Covers

UL 489 requires that some markings be visible without removing trims or covers. This location is typically referred to as the handle escutcheon (see photo 1).

Markings Visible with Trims or Covers Removed

UL 489 requires other markings be visible on an installed circuit breaker with trims or covers removed. This location is typically referred to as the face of the circuit breaker (see photos 2–4).

Other markings which should be visible with trims or covers removed are:

Independent trip – Multi-pole circuit breakers are constructed with either a common trip, where all poles are mechanically tripped when one of the poles trips, or an independent trip construction where only the pole that is involved with the overcurrent condition trips. If a 2-pole circuit breaker does not have an internal com-

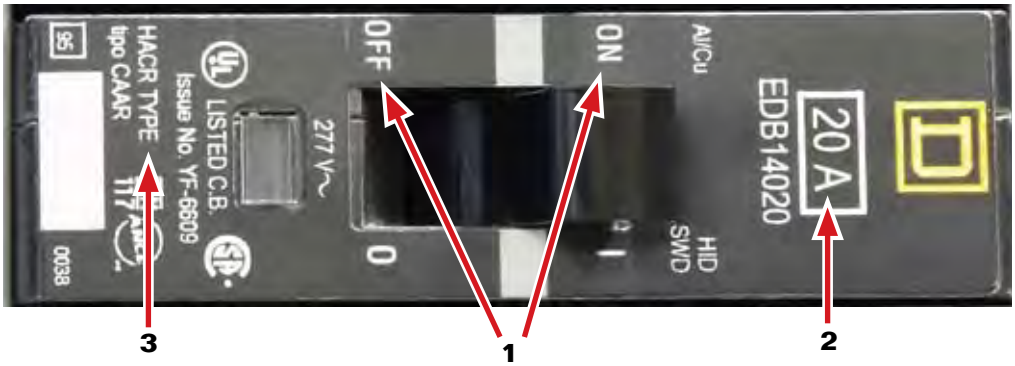
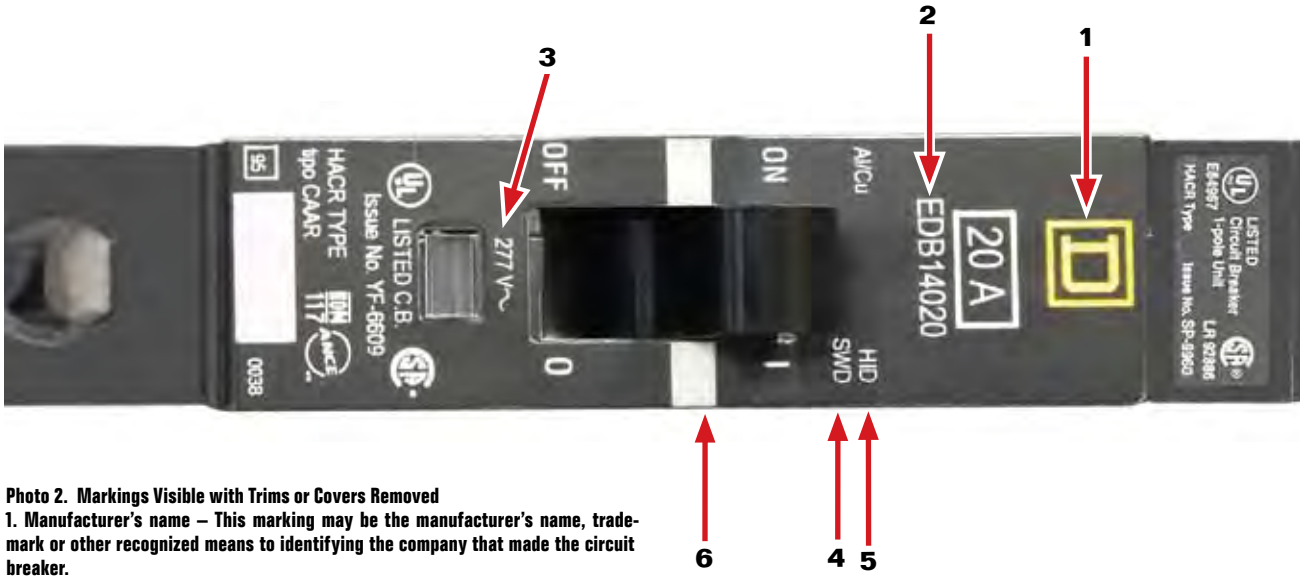


Photo 1. Markings Visible with Trims or Covers in Place

1. ON and OFF – The ON and OFF (closed and open) positions of the handle must be marked (*NEC* 240.81). These positions may also be marked with the internationally recognized "I" and "O" symbols, although this is not a UL requirement. If these markings are not visible when a motor operator is installed over the circuit breaker markings, then they must appear on the motor operator. Motor operators may be found in applications where remote or automatic operation of a circuit breaker is required.

2. Ampere rating (if 100 A or less) – The ampere rating may be located on the handle escutcheon or on the handle itself (*NEC* 240.83(B)). Circuit breakers that are rated more than 100 A may have their ampere rating marked in a position that is not visible with trims or covers in place.



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1. Manufacturer's name – This marking may be the manufacturer's name, trademark or other recognized means to identifying the company that made the circuit breaker.

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3. Voltage rating – All circuit breakers must be marked with a voltage rating. If the rating is not marked "ac" or "dc," then it is suitable for both. 120/240 V rated circuit breakers are suitable for use on single and three-phase 4-wire systems where the line-to-ground voltage does not exceed 120 V. Wye rated circuit breakers such as those rated 480Y/277 V, are suitable for use on three-phase 4-wire systems where the voltage to ground does not exceed 277 V. Special attention needs to be given to high leg or corner-grounded delta systems to insure that the circuit breaker has the appropriate rating. A review to see that the circuit breakers installed have a voltage rating suitable for the application is paramount for a code-compliant installation (*NEC* 240.83(E)).

4. SWD – 15- or 20-A circuit breakers rated 347 V or less may be marked "SWD," meaning that they are suitable for switching fluorescent lighting loads on a regular basis (*NEC* 240.83(D)). These circuit breakers are evaluated for high endurance use, since they will be used similar to a light switch.

5. HID – 50 A or less circuit breakers rated 480 V or less may be marked "HID," meaning they are suitable for switching high intensity discharge or fluorescent lighting loads on a regular basis. These circuit breakers may employ a different construction than a standard SWD circuit breaker in order to address the high inrush current resulting from the lower power factor created by the HID lighting (*NEC* 240.83(D)). These circuit breakers also undergo additional endurance evaluation to demonstrate their ability to perform the switching duty.

6. Trip and reset – Circuit breaker handles typically assume an intermediate position when tripped. This position must either be marked on the circuit breaker or on the equipment into which it is to be installed. If these markings are not visible when a motor operator is installed, then a "tripped" marking may appear on the motor operator.

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Markings Found in Other Locations

The markings we will discuss below may appear in any location except the back of the circuit breaker. These markings include:

40°C – This marking indicates the maximum ambient temperature in which the circuit breaker can be applied at its marked ampere rating without rerating the ampacity of the circuit breaker. This marking is required for thermal-magnetic circuit breakers and is optional for electronic trip circuit breakers unless they are only suitable for a 25°C ambient, in which case they must be marked 25°C. When the ambient temperature rises above 40°C, the designer may need to consult the manufacturer to obtain rerating information (see item 4 in photo 3).

Class CTL – Circuit breakers marked Class CTL have a rejection means designed into the circuit breaker. Class CTL panelboards or assemblies, in conjunction with Class CTL circuit breakers, prevent more circuit breaker poles from being installed than the number for which the equipment is rated.

Photo 3. Markings Visible with Trims or Covers Removed

1. Line and load designation – Circuit breakers marked with "line" and "load" designations are not suitable for reverse connection. Circuit breakers with interchangeable trip units must be marked "line" and "load" unless there is no risk of shock when changing the trip unit.

2. Interrupting ratings – All circuit breakers with an interrupting rating more than 5000 A must be marked with an interrupting rating (*NEC* 240.83(C)). Interrupting ratings are stated in RMS symmetrical amperes. If the short-circuit current rating of the equipment in which the circuit breaker is installed is less than the interrupting rating of the circuit breaker, then the lesser rating applies. Circuit breakers should be reviewed after installation to ensure they have an interrupting rating suitable for the application. This marking may be found in any location except the back of circuit breakers that are 1-½ inches wide per pole or less due to the size constraint.

3. Ampere rating (if more than 100 A) – The ampere rating of a circuit breaker larger than 100 A may be found in a location that is visible after the cover or trim is removed. This marking requirement also applies to interchangeable trip units (*NEC* 240.83(A)).

5. Terminations (Cu-Al wire) – Circuit breakers must be marked with the type material (Cu-Al) and size of wire for which their terminals are suitable for use. If suitable for use with only copper or only aluminum, then the word "only" must be used. The abbreviations "CU" and "AL" are generally found on circuit breakers as permitted by the product standard. If only solid 10–14 AWG wire can be used, then that information must be noted. The number of wires per terminal will also be noted if more than one wire per terminal is permitted (*NEC* 110.14(A)).

6. Tightening torque – The nominal torque for all field-wiring terminals must be marked. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be found in any location except on the back.

7. Wire temperature ratings – Circuit breakers rated 125 A or less may be marked as suitable for use with 60°C, 60/75°C or 75°C only wire. Circuit breakers rated more than 125 A are rated for use with 75°C wire; the marking is optional. It is always permissible to use wire with a higher temperature rating, but it must be sized in accordance with the temperature marking on the circuit breaker and *NEC* Table 310.16. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be in any location except on the back (*NEC* 110.14(C)1).

8. HACR type – This marking indicates the circuit breaker is suitable for use with the group motor installations typically found in heating, air conditioning and refrigeration equipment. The *NEC* 2005 no longer has this marking requirement. The electrical industry determined that circuit breakers are considered suitable for use with such equipment without any further testing, therefore, the HACR marking is no longer required on air conditioning and refrigeration equipment or on circuit breakers for use in these applications. The requirement for this marking has also been removed from the UL 1995 product standard for HVAC equipment (see item 3 in photo 1).

Maximum wire size – Circuit breakers are typically marked with a wire range, however that marking is not mandatory. If the circuit breaker cannot accept the next larger wire size required for the ampere rating, then the maximum wire size must be marked in any location except the back (see item 5 in photo 3).

Separately shipped connectors –If connectors are not factory installed on a circuit breaker, then it must be marked with the proper connectors or terminal kits required in any location except the back (see item 8 in photo 3).

Ground-Fault Protection for People

The GFCI function, as part of a circuit breaker, provides ground-fault protection for people and has a number of unique marking and instruction requirements.

Test function – The GFCI has a test function that requires action upon installation and on a monthly basis. GFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate monthly testing.

Photo 4. 1Ø – 3Ø Marking
1. 1Ø – 3Ø marking – A 2-pole circuit breaker used to protect a 3-phase load on a corner-grounded delta system must be rated and marked for such an installation. Circuit breakers marked "1-phase – 3-phase" or "1Ø – 3Ø "are suitable for use on 3-phase corner-grounded delta or single-phase circuits (*NEC* 240.85).

Suitable for continuous operation at 100% of ampere rating only if used in a recognized enclosure 52 in. (1321 mm) by 20 in. (508 mm) x 7 1/2 in. (190 mm) deep or larger. Use only with 90° rated wire sized per the ampacity of 75° rated wire.

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Trip level – GFPE circuit breakers must be marked with their trip threshold in milliamperes in a location accessible without removing trims or covers.

Instructions – All GFPE circuit breakers must include instructions for the installer.

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Device identification – AFCIs must also be identified appropriately. Branch/feeder or Combination type AFCIs must be so marked in a location visible when the trims or covers are removed. This is an important marking to note as we move into 2008, as *NEC*-2005 requires Combination AFCIs in bedrooms effective January 1, 2008 (*NEC* 210.12).

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UL 489 requires that some markings be visible without removing trims or covers. This location is typically referred to as the handle escutcheon (see photo 1).

Markings Visible with Trims or Covers Removed

UL 489 requires other markings be visible on an installed circuit breaker with trims or covers removed. This location is typically referred to as the face of the circuit breaker (see photos 2–4).

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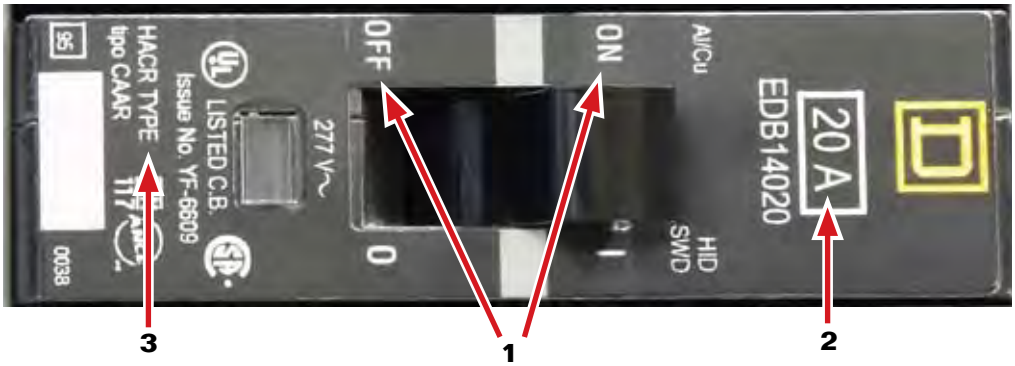


Photo 1. Markings Visible with Trims or Covers in Place

1. ON and OFF – The ON and OFF (closed and open) positions of the handle must be marked (*NEC* 240.81). These positions may also be marked with the internationally recognized "I" and "O" symbols, although this is not a UL requirement. If these markings are not visible when a motor operator is installed over the circuit breaker markings, then they must appear on the motor operator. Motor operators may be found in applications where remote or automatic operation of a circuit breaker is required.

2. Ampere rating (if 100 A or less) – The ampere rating may be located on the handle escutcheon or on the handle itself (*NEC* 240.83(B)). Circuit breakers that are rated more than 100 A may have their ampere rating marked in a position that is not visible with trims or covers in place.

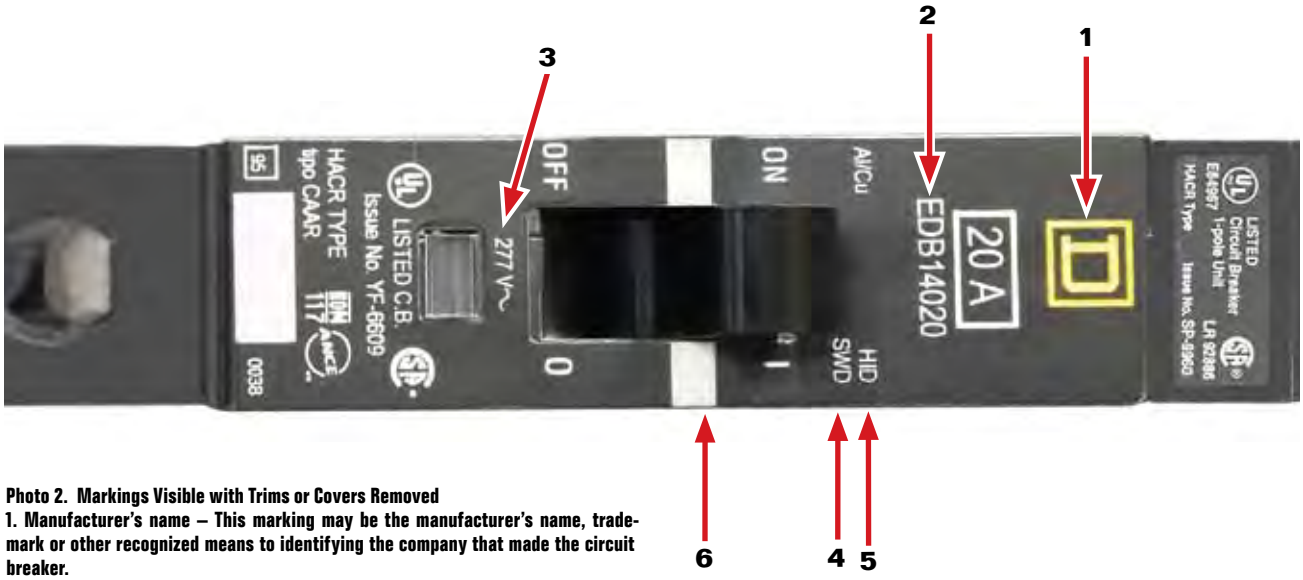


Photo 2. Markings Visible with Trims or Covers Removed

1. Manufacturer's name – This marking may be the manufacturer's name, trademark or other recognized means to identifying the company that made the circuit breaker.

2. Type designation – All circuit breakers are marked with a type designation, which may be a catalog number prefix or a separate designation. Equipment labels, such as on panelboards, will list the circuit breaker types suitable for use. Note that the word "type" may or may not be used on the circuit breaker or equipment labels. It is important to review the markings on the equipment, such as a panelboard, to make sure the circuit breaker designations on the equipment match the marking on the circuit breaker.

3. Voltage rating – All circuit breakers must be marked with a voltage rating. If the rating is not marked "ac" or "dc," then it is suitable for both. 120/240 V rated circuit breakers are suitable for use on single and three-phase 4-wire systems where the line-to-ground voltage does not exceed 120 V. Wye rated circuit breakers such as those rated 480Y/277 V, are suitable for use on three-phase 4-wire systems where the voltage to ground does not exceed 277 V. Special attention needs to be given to high leg or corner-grounded delta systems to insure that the circuit breaker has the appropriate rating. A review to see that the circuit breakers installed have a voltage rating suitable for the application is paramount for a code-compliant installation (*NEC* 240.83(E)).

4. SWD – 15- or 20-A circuit breakers rated 347 V or less may be marked "SWD," meaning that they are suitable for switching fluorescent lighting loads on a regular basis (*NEC* 240.83(D)). These circuit breakers are evaluated for high endurance use, since they will be used similar to a light switch.

5. HID – 50 A or less circuit breakers rated 480 V or less may be marked "HID," meaning they are suitable for switching high intensity discharge or fluorescent lighting loads on a regular basis. These circuit breakers may employ a different construction than a standard SWD circuit breaker in order to address the high inrush current resulting from the lower power factor created by the HID lighting (*NEC* 240.83(D)). These circuit breakers also undergo additional endurance evaluation to demonstrate their ability to perform the switching duty.

6. Trip and reset – Circuit breaker handles typically assume an intermediate position when tripped. This position must either be marked on the circuit breaker or on the equipment into which it is to be installed. If these markings are not visible when a motor operator is installed, then a "tripped" marking may appear on the motor operator.

mon trip feature, then it must be marked "Independent Trip" or "No Common Trip." *NEC* 240.20(B) is the foundational requirement for a common trip function in a circuit breaker; however, it also goes on to explain where independent trip is permitted.

For Replacement Use Only not-CTL – The Class CTL (circuit limiting) panelboard has only been in existence for about 25 years, even though the lighting and appliance branch circuit panelboard has been in the *NEC* for decades. CTL panelboards have a rejection means designed to reject more than the appropriate number of circuit breakers that can be installed in the panel. The marking "For replacement Use Only Not CTL Assemblies" means that the circuit breaker does not have CTL rejection provisions and is intended for replacement in older equipment pre-dating the CTL requirements for circuit breakers and panelboards. Circuit breakers with this designation should not be installed in a panelboard marked "Class CTL Panelboard" since that would be a violation of the listing of the assembly [*NEC* 110.3(B)].

Markings Found in Other Locations

The markings we will discuss below may appear in any location except the back of the circuit breaker. These markings include:

40°C – This marking indicates the maximum ambient temperature in which the circuit breaker can be applied at its marked ampere rating without rerating the ampacity of the circuit breaker. This marking is required for thermal-magnetic circuit breakers and is optional for electronic trip circuit breakers unless they are only suitable for a 25°C ambient, in which case they must be marked 25°C. When the ambient temperature rises above 40°C, the designer may need to consult the manufacturer to obtain rerating information (see item 4 in photo 3).

Class CTL – Circuit breakers marked Class CTL have a rejection means designed into the circuit breaker. Class CTL panelboards or assemblies, in conjunction with Class CTL circuit breakers, prevent more circuit breaker poles from being installed than the number for which the equipment is rated.

Photo 3. Markings Visible with Trims or Covers Removed

1. Line and load designation – Circuit breakers marked with "line" and "load" designations are not suitable for reverse connection. Circuit breakers with interchangeable trip units must be marked "line" and "load" unless there is no risk of shock when changing the trip unit.

2. Interrupting ratings – All circuit breakers with an interrupting rating more than 5000 A must be marked with an interrupting rating (*NEC* 240.83(C)). Interrupting ratings are stated in RMS symmetrical amperes. If the short-circuit current rating of the equipment in which the circuit breaker is installed is less than the interrupting rating of the circuit breaker, then the lesser rating applies. Circuit breakers should be reviewed after installation to ensure they have an interrupting rating suitable for the application. This marking may be found in any location except the back of circuit breakers that are 1-½ inches wide per pole or less due to the size constraint.

3. Ampere rating (if more than 100 A) – The ampere rating of a circuit breaker larger than 100 A may be found in a location that is visible after the cover or trim is removed. This marking requirement also applies to interchangeable trip units (*NEC* 240.83(A)).

5. Terminations (Cu-Al wire) – Circuit breakers must be marked with the type material (Cu-Al) and size of wire for which their terminals are suitable for use. If suitable for use with only copper or only aluminum, then the word "only" must be used. The abbreviations "CU" and "AL" are generally found on circuit breakers as permitted by the product standard. If only solid 10–14 AWG wire can be used, then that information must be noted. The number of wires per terminal will also be noted if more than one wire per terminal is permitted (*NEC* 110.14(A)).

6. Tightening torque – The nominal torque for all field-wiring terminals must be marked. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be found in any location except on the back.

7. Wire temperature ratings – Circuit breakers rated 125 A or less may be marked as suitable for use with 60°C, 60/75°C or 75°C only wire. Circuit breakers rated more than 125 A are rated for use with 75°C wire; the marking is optional. It is always permissible to use wire with a higher temperature rating, but it must be sized in accordance with the temperature marking on the circuit breaker and *NEC* Table 310.16. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be in any location except on the back (*NEC* 110.14(C)1).

8. HACR type – This marking indicates the circuit breaker is suitable for use with the group motor installations typically found in heating, air conditioning and refrigeration equipment. The *NEC* 2005 no longer has this marking requirement. The electrical industry determined that circuit breakers are considered suitable for use with such equipment without any further testing, therefore, the HACR marking is no longer required on air conditioning and refrigeration equipment or on circuit breakers for use in these applications. The requirement for this marking has also been removed from the UL 1995 product standard for HVAC equipment (see item 3 in photo 1).

Maximum wire size – Circuit breakers are typically marked with a wire range, however that marking is not mandatory. If the circuit breaker cannot accept the next larger wire size required for the ampere rating, then the maximum wire size must be marked in any location except the back (see item 5 in photo 3).

Separately shipped connectors –If connectors are not factory installed on a circuit breaker, then it must be marked with the proper connectors or terminal kits required in any location except the back (see item 8 in photo 3).

Ground-Fault Protection for People

The GFCI function, as part of a circuit breaker, provides ground-fault protection for people and has a number of unique marking and instruction requirements.

Test function – The GFCI has a test function that requires action upon installation and on a monthly basis. GFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate monthly testing.

Photo 4. 1Ø – 3Ø Marking
1. 1Ø – 3Ø marking – A 2-pole circuit breaker used to protect a 3-phase load on a corner-grounded delta system must be rated and marked for such an installation. Circuit breakers marked "1-phase – 3-phase" or "1Ø – 3Ø "are suitable for use on 3-phase corner-grounded delta or single-phase circuits (*NEC* 240.85).

Suitable for continuous operation at 100% of ampere rating only if used in a recognized enclosure 52 in. (1321 mm) by 20 in. (508 mm) x 7 1/2 in. (190 mm) deep or larger. Use only with 90° rated wire sized per the ampacity of 75° rated wire.

Figure 1. 100% Rated Marking
100 percent continuous rated – Circuit breakers are typically intended for use at not more than 80% of rated current where the load is considered continuous, or will continue for 3 hours or more (*NEC* 210.20). However, some circuit breakers are rated for continuous use at 100% of their current rating. These circuit breakers must be so marked in any location except on the back. Enclosure information such as a specific type or specific volume must also be marked. A requirement for the use of 90°C insulated wire sized to the 75°C column in *NEC* Table 310.16 and specific ventilation requirements may also be marked on the circuit breaker or equipment (*NEC* 210.20(A) and 215.3).

"Class A" marking – A "Class A" ground-fault device is intended to protect people. The Class A marking indicates that the trip threshold of the GFCI is between 4 mA and 6 mA. This marking may be in any location except the back.

Instructions – All GFCI circuit breakers must include instructions for the installer plus instructions on the use of the test function. A hangtag or self-adhesive label must also be provided, instructing the user to test the GFCI at least monthly. Inspectors should check to see that the tag or label has been properly installed.

Ground-Fault Protection for Equipment

Circuit breakers may also include a ground-fault protection for equipment (GFPE) function that, like GFCIs, has a number of unique marking and instruction requirements.

Test function – GFPE circuit breakers may have a test button or switch that may be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Trip level – GFPE circuit breakers must be marked with their trip threshold in milliamperes in a location accessible without removing trims or covers.

Instructions – All GFPE circuit breakers must include instructions for the installer.

Arc-Fault Protection

Circuit breakers may also include arc-fault protection (AFCI) that, like GFCIs, also has a number of unique marking and instruction requirements.

Device identification – AFCIs must also be identified appropriately. Branch/feeder or Combination type AFCIs must be so marked in a location visible when the trims or covers are removed. This is an important marking to note as we move into 2008, as *NEC*-2005 requires Combination AFCIs in bedrooms effective January 1, 2008 (*NEC* 210.12).

Test function – AFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Instructions – All AFCI circuit breakers must include instructions for the installer.

Circuit Breaker Markings Ensure a Safe Electrical Installation

So why are all of these markings on circuit breakers? Without them, it would be nearly impossible to install or inspect an installation for the appropriate performance ratings and fundamental electrical connections. When designing or completing an installation, key items to review are:

1. Are the voltage, continuous current, and interrupting ratings appropriate for the application?
2. Does the application require SWD or HID ratings?
3. Is the wire type and size appropriate for the circuit breaker?
4. Is the circuit breaker suitable for the equipment in which it is installed? Have other protective functions such as GFCI or AFCI been provided as required by the *NEC*?
5. Is the temperature rating of the circuit breaker suitable for the application?

The UL Marking Guide for Molded Case Circuit Breakers is a valuable resource to understand circuit breaker markings that may further explain these and other markings in detail. If you have questions about CB markings not answered here, consult the Marking Guide or the manufacturer to assist in an *NEC*-compliant installation.

Mr. Larsen, P. E., is the manager, Industry Standards for Circuit Protection for Square D Company / Schneider Electric. Mr. Larsen has served in various positions over the past 32 years. He is a member of several UL, IEEE and NEMA technical committees. His other memberships include NFPA, IAEI and CANENA.

What Do the Markings on Circuit Breakers Mean?

by Ed Larsen, P. E.

Have you ever been confused about what the markings on circuit breakers mean? Understanding the markings on electrical equipment is a fundamental need to ensure a safe and reliable electrical installation. Circuit breaker marking requirements are established by the requirements found in the *NEC* and the UL 489 product standard. This article will discuss the most common markings and where they can be found.

The UL 489 product standard for Molded Case Circuit Breakers specifies the information to be marked on circuit breakers and where it is to be located, so let's discuss what information needs to be marked on the circuit breaker and the location where you will find those markings. Keep in mind the UL® standard specifies minimum requirements. Circuit breaker manufacturers may provide additional information or provide information in a more convenient location.

Markings Visible without Removing Trims or Covers

UL 489 requires that some markings be visible without removing trims or covers. This location is typically referred to as the handle escutcheon (see photo 1).

Markings Visible with Trims or Covers Removed

UL 489 requires other markings be visible on an installed circuit breaker with trims or covers removed. This location is typically referred to as the face of the circuit breaker (see photos 2–4).

Other markings which should be visible with trims or covers removed are:

Independent trip – Multi-pole circuit breakers are constructed with either a common trip, where all poles are mechanically tripped when one of the poles trips, or an independent trip construction where only the pole that is involved with the overcurrent condition trips. If a 2-pole circuit breaker does not have an internal com-

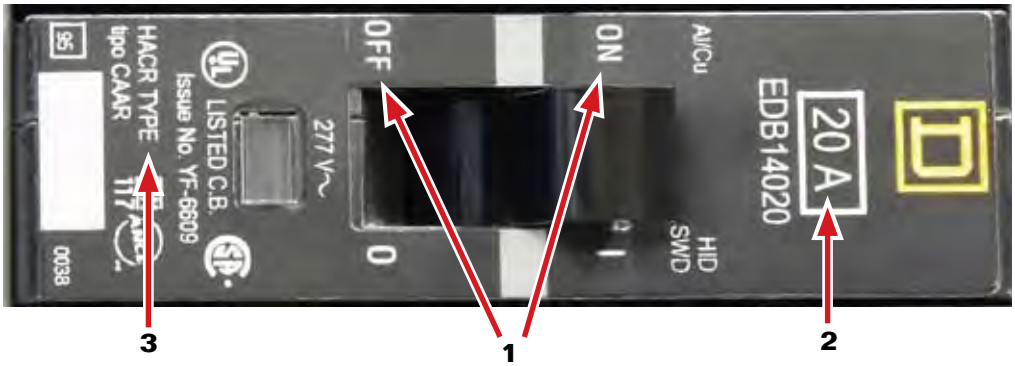
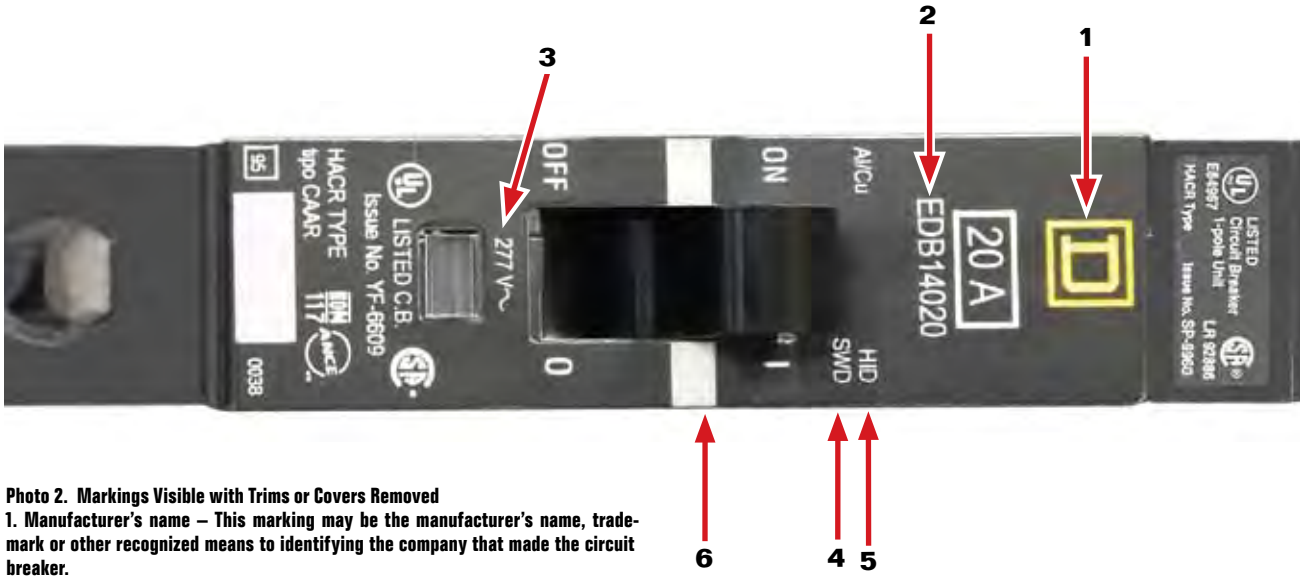


Photo 1. Markings Visible with Trims or Covers in Place

1. ON and OFF – The ON and OFF (closed and open) positions of the handle must be marked (*NEC* 240.81). These positions may also be marked with the internationally recognized "I" and "O" symbols, although this is not a UL requirement. If these markings are not visible when a motor operator is installed over the circuit breaker markings, then they must appear on the motor operator. Motor operators may be found in applications where remote or automatic operation of a circuit breaker is required.

2. Ampere rating (if 100 A or less) – The ampere rating may be located on the handle escutcheon or on the handle itself (*NEC* 240.83(B)). Circuit breakers that are rated more than 100 A may have their ampere rating marked in a position that is not visible with trims or covers in place.



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1. Manufacturer's name – This marking may be the manufacturer's name, trademark or other recognized means to identifying the company that made the circuit breaker.

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3. Voltage rating – All circuit breakers must be marked with a voltage rating. If the rating is not marked "ac" or "dc," then it is suitable for both. 120/240 V rated circuit breakers are suitable for use on single and three-phase 4-wire systems where the line-to-ground voltage does not exceed 120 V. Wye rated circuit breakers such as those rated 480Y/277 V, are suitable for use on three-phase 4-wire systems where the voltage to ground does not exceed 277 V. Special attention needs to be given to high leg or corner-grounded delta systems to insure that the circuit breaker has the appropriate rating. A review to see that the circuit breakers installed have a voltage rating suitable for the application is paramount for a code-compliant installation (*NEC* 240.83(E)).

4. SWD – 15- or 20-A circuit breakers rated 347 V or less may be marked "SWD," meaning that they are suitable for switching fluorescent lighting loads on a regular basis (*NEC* 240.83(D)). These circuit breakers are evaluated for high endurance use, since they will be used similar to a light switch.

5. HID – 50 A or less circuit breakers rated 480 V or less may be marked "HID," meaning they are suitable for switching high intensity discharge or fluorescent lighting loads on a regular basis. These circuit breakers may employ a different construction than a standard SWD circuit breaker in order to address the high inrush current resulting from the lower power factor created by the HID lighting (*NEC* 240.83(D)). These circuit breakers also undergo additional endurance evaluation to demonstrate their ability to perform the switching duty.

6. Trip and reset – Circuit breaker handles typically assume an intermediate position when tripped. This position must either be marked on the circuit breaker or on the equipment into which it is to be installed. If these markings are not visible when a motor operator is installed, then a "tripped" marking may appear on the motor operator.

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Markings Found in Other Locations

The markings we will discuss below may appear in any location except the back of the circuit breaker. These markings include:

40°C – This marking indicates the maximum ambient temperature in which the circuit breaker can be applied at its marked ampere rating without rerating the ampacity of the circuit breaker. This marking is required for thermal-magnetic circuit breakers and is optional for electronic trip circuit breakers unless they are only suitable for a 25°C ambient, in which case they must be marked 25°C. When the ambient temperature rises above 40°C, the designer may need to consult the manufacturer to obtain rerating information (see item 4 in photo 3).

Class CTL – Circuit breakers marked Class CTL have a rejection means designed into the circuit breaker. Class CTL panelboards or assemblies, in conjunction with Class CTL circuit breakers, prevent more circuit breaker poles from being installed than the number for which the equipment is rated.

Photo 3. Markings Visible with Trims or Covers Removed

1. Line and load designation – Circuit breakers marked with "line" and "load" designations are not suitable for reverse connection. Circuit breakers with interchangeable trip units must be marked "line" and "load" unless there is no risk of shock when changing the trip unit.

2. Interrupting ratings – All circuit breakers with an interrupting rating more than 5000 A must be marked with an interrupting rating (*NEC* 240.83(C)). Interrupting ratings are stated in RMS symmetrical amperes. If the short-circuit current rating of the equipment in which the circuit breaker is installed is less than the interrupting rating of the circuit breaker, then the lesser rating applies. Circuit breakers should be reviewed after installation to ensure they have an interrupting rating suitable for the application. This marking may be found in any location except the back of circuit breakers that are 1-½ inches wide per pole or less due to the size constraint.

3. Ampere rating (if more than 100 A) – The ampere rating of a circuit breaker larger than 100 A may be found in a location that is visible after the cover or trim is removed. This marking requirement also applies to interchangeable trip units (*NEC* 240.83(A)).

5. Terminations (Cu-Al wire) – Circuit breakers must be marked with the type material (Cu-Al) and size of wire for which their terminals are suitable for use. If suitable for use with only copper or only aluminum, then the word "only" must be used. The abbreviations "CU" and "AL" are generally found on circuit breakers as permitted by the product standard. If only solid 10–14 AWG wire can be used, then that information must be noted. The number of wires per terminal will also be noted if more than one wire per terminal is permitted (*NEC* 110.14(A)).

6. Tightening torque – The nominal torque for all field-wiring terminals must be marked. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be found in any location except on the back.

7. Wire temperature ratings – Circuit breakers rated 125 A or less may be marked as suitable for use with 60°C, 60/75°C or 75°C only wire. Circuit breakers rated more than 125 A are rated for use with 75°C wire; the marking is optional. It is always permissible to use wire with a higher temperature rating, but it must be sized in accordance with the temperature marking on the circuit breaker and *NEC* Table 310.16. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be in any location except on the back (*NEC* 110.14(C)1).

8. HACR type – This marking indicates the circuit breaker is suitable for use with the group motor installations typically found in heating, air conditioning and refrigeration equipment. The *NEC* 2005 no longer has this marking requirement. The electrical industry determined that circuit breakers are considered suitable for use with such equipment without any further testing, therefore, the HACR marking is no longer required on air conditioning and refrigeration equipment or on circuit breakers for use in these applications. The requirement for this marking has also been removed from the UL 1995 product standard for HVAC equipment (see item 3 in photo 1).

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Separately shipped connectors –If connectors are not factory installed on a circuit breaker, then it must be marked with the proper connectors or terminal kits required in any location except the back (see item 8 in photo 3).

Ground-Fault Protection for People

The GFCI function, as part of a circuit breaker, provides ground-fault protection for people and has a number of unique marking and instruction requirements.

Test function – The GFCI has a test function that requires action upon installation and on a monthly basis. GFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate monthly testing.

Photo 4. 1Ø – 3Ø Marking
1. 1Ø – 3Ø marking – A 2-pole circuit breaker used to protect a 3-phase load on a corner-grounded delta system must be rated and marked for such an installation. Circuit breakers marked "1-phase – 3-phase" or "1Ø – 3Ø "are suitable for use on 3-phase corner-grounded delta or single-phase circuits (*NEC* 240.85).

Suitable for continuous operation at 100% of ampere rating only if used in a recognized enclosure 52 in. (1321 mm) by 20 in. (508 mm) x 7 1/2 in. (190 mm) deep or larger. Use only with 90° rated wire sized per the ampacity of 75° rated wire.

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100 percent continuous rated – Circuit breakers are typically intended for use at not more than 80% of rated current where the load is considered continuous, or will continue for 3 hours or more (*NEC* 210.20). However, some circuit breakers are rated for continuous use at 100% of their current rating. These circuit breakers must be so marked in any location except on the back. Enclosure information such as a specific type or specific volume must also be marked. A requirement for the use of 90°C insulated wire sized to the 75°C column in *NEC* Table 310.16 and specific ventilation requirements may also be marked on the circuit breaker or equipment (*NEC* 210.20(A) and 215.3).

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Test function – GFPE circuit breakers may have a test button or switch that may be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Trip level – GFPE circuit breakers must be marked with their trip threshold in milliamperes in a location accessible without removing trims or covers.

Instructions – All GFPE circuit breakers must include instructions for the installer.

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Device identification – AFCIs must also be identified appropriately. Branch/feeder or Combination type AFCIs must be so marked in a location visible when the trims or covers are removed. This is an important marking to note as we move into 2008, as *NEC*-2005 requires Combination AFCIs in bedrooms effective January 1, 2008 (*NEC* 210.12).

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Markings Visible without Removing Trims or Covers

UL 489 requires that some markings be visible without removing trims or covers. This location is typically referred to as the handle escutcheon (see photo 1).

Markings Visible with Trims or Covers Removed

UL 489 requires other markings be visible on an installed circuit breaker with trims or covers removed. This location is typically referred to as the face of the circuit breaker (see photos 2–4).

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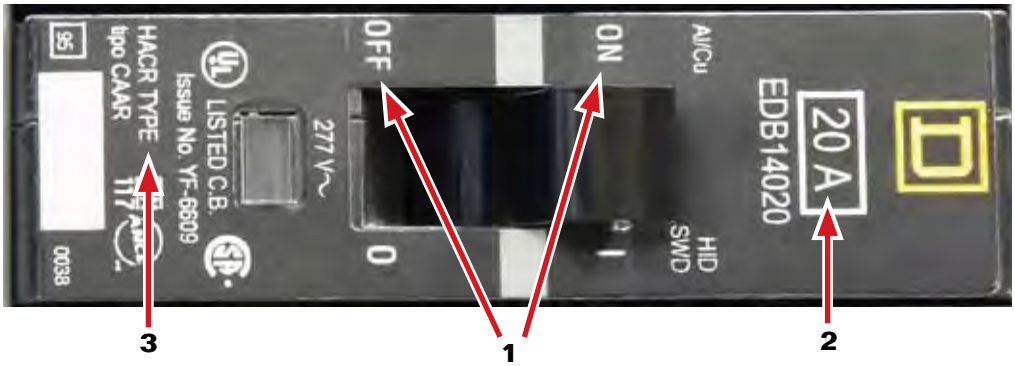

Photo 1. Markings Visible with Trims or Covers in Place

1. ON and OFF – The ON and OFF (closed and open) positions of the handle must be marked (*NEC* 240.81). These positions may also be marked with the internationally recognized "I" and "O" symbols, although this is not a UL requirement. If these markings are not visible when a motor operator is installed over the circuit breaker markings, then they must appear on the motor operator. Motor operators may be found in applications where remote or automatic operation of a circuit breaker is required.

2. Ampere rating (if 100 A or less) – The ampere rating may be located on the handle escutcheon or on the handle itself (*NEC* 240.83(B)). Circuit breakers that are rated more than 100 A may have their ampere rating marked in a position that is not visible with trims or covers in place.

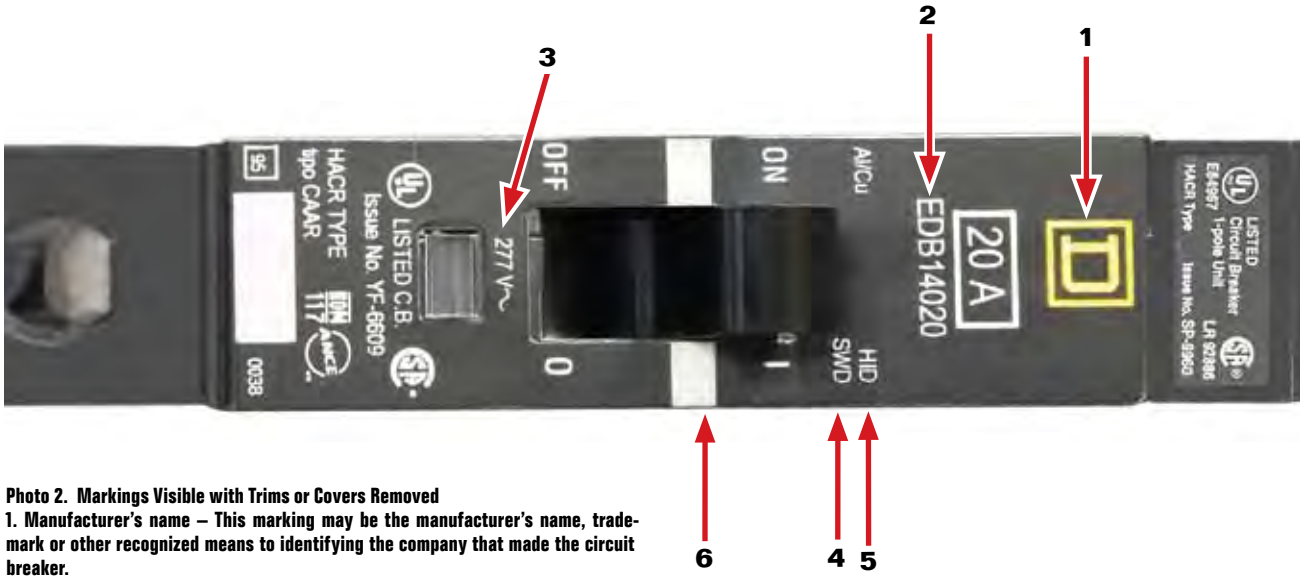

Photo 2. Markings Visible with Trims or Covers Removed

1. Manufacturer's name – This marking may be the manufacturer's name, trademark or other recognized means to identifying the company that made the circuit breaker.

2. Type designation – All circuit breakers are marked with a type designation, which may be a catalog number prefix or a separate designation. Equipment labels, such as on panelboards, will list the circuit breaker types suitable for use. Note that the word "type" may or may not be used on the circuit breaker or equipment labels. It is important to review the markings on the equipment, such as a panelboard, to make sure the circuit breaker designations on the equipment match the marking on the circuit breaker.

3. Voltage rating – All circuit breakers must be marked with a voltage rating. If the rating is not marked "ac" or "dc," then it is suitable for both. 120/240 V rated circuit breakers are suitable for use on single and three-phase 4-wire systems where the line-to-ground voltage does not exceed 120 V. Wye rated circuit breakers such as those rated 480Y/277 V, are suitable for use on three-phase 4-wire systems where the voltage to ground does not exceed 277 V. Special attention needs to be given to high leg or corner-grounded delta systems to insure that the circuit breaker has the appropriate rating. A review to see that the circuit breakers installed have a voltage rating suitable for the application is paramount for a code-compliant installation (*NEC* 240.83(E)).

4. SWD – 15- or 20-A circuit breakers rated 347 V or less may be marked "SWD," meaning that they are suitable for switching fluorescent lighting loads on a regular basis (*NEC* 240.83(D)). These circuit breakers are evaluated for high endurance use, since they will be used similar to a light switch.

5. HID – 50 A or less circuit breakers rated 480 V or less may be marked "HID," meaning they are suitable for switching high intensity discharge or fluorescent lighting loads on a regular basis. These circuit breakers may employ a different construction than a standard SWD circuit breaker in order to address the high inrush current resulting from the lower power factor created by the HID lighting (*NEC* 240.83(D)). These circuit breakers also undergo additional endurance evaluation to demonstrate their ability to perform the switching duty.

6. Trip and reset – Circuit breaker handles typically assume an intermediate position when tripped. This position must either be marked on the circuit breaker or on the equipment into which it is to be installed. If these markings are not visible when a motor operator is installed, then a "tripped" marking may appear on the motor operator.

mon trip feature, then it must be marked "Independent Trip" or "No Common Trip." *NEC* 240.20(B) is the foundational requirement for a common trip function in a circuit breaker; however, it also goes on to explain where independent trip is permitted.

For Replacement Use Only not-CTL – The Class CTL (circuit limiting) panelboard has only been in existence for about 25 years, even though the lighting and appliance branch circuit panelboard has been in the *NEC* for decades. CTL panelboards have a rejection means designed to reject more than the appropriate number of circuit breakers that can be installed in the panel. The marking "For replacement Use Only Not CTL Assemblies" means that the circuit breaker does not have CTL rejection provisions and is intended for replacement in older equipment pre-dating the CTL requirements for circuit breakers and panelboards. Circuit breakers with this designation should not be installed in a panelboard marked "Class CTL Panelboard" since that would be a violation of the listing of the assembly [*NEC* 110.3(B)].

Markings Found in Other Locations

The markings we will discuss below may appear in any location except the back of the circuit breaker. These markings include:

40°C – This marking indicates the maximum ambient temperature in which the circuit breaker can be applied at its marked ampere rating without rerating the ampacity of the circuit breaker. This marking is required for thermal-magnetic circuit breakers and is optional for electronic trip circuit breakers unless they are only suitable for a 25°C ambient, in which case they must be marked 25°C. When the ambient temperature rises above 40°C, the designer may need to consult the manufacturer to obtain rerating information (see item 4 in photo 3).

Class CTL – Circuit breakers marked Class CTL have a rejection means designed into the circuit breaker. Class CTL panelboards or assemblies, in conjunction with Class CTL circuit breakers, prevent more circuit breaker poles from being installed than the number for which the equipment is rated.

Photo 3. Markings Visible with Trims or Covers Removed

1. Line and load designation – Circuit breakers marked with "line" and "load" designations are not suitable for reverse connection. Circuit breakers with interchangeable trip units must be marked "line" and "load" unless there is no risk of shock when changing the trip unit.

2. Interrupting ratings – All circuit breakers with an interrupting rating more than 5000 A must be marked with an interrupting rating (*NEC* 240.83(C)). Interrupting ratings are stated in RMS symmetrical amperes. If the short-circuit current rating of the equipment in which the circuit breaker is installed is less than the interrupting rating of the circuit breaker, then the lesser rating applies. Circuit breakers should be reviewed after installation to ensure they have an interrupting rating suitable for the application. This marking may be found in any location except the back of circuit breakers that are 1-½ inches wide per pole or less due to the size constraint.

3. Ampere rating (if more than 100 A) – The ampere rating of a circuit breaker larger than 100 A may be found in a location that is visible after the cover or trim is removed. This marking requirement also applies to interchangeable trip units (*NEC* 240.83(A)).

5. Terminations (Cu-Al wire) – Circuit breakers must be marked with the type material (Cu-Al) and size of wire for which their terminals are suitable for use. If suitable for use with only copper or only aluminum, then the word "only" must be used. The abbreviations "CU" and "AL" are generally found on circuit breakers as permitted by the product standard. If only solid 10–14 AWG wire can be used, then that information must be noted. The number of wires per terminal will also be noted if more than one wire per terminal is permitted (*NEC* 110.14(A)).

6. Tightening torque – The nominal torque for all field-wiring terminals must be marked. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be found in any location except on the back.

7. Wire temperature ratings – Circuit breakers rated 125 A or less may be marked as suitable for use with 60°C, 60/75°C or 75°C only wire. Circuit breakers rated more than 125 A are rated for use with 75°C wire; the marking is optional. It is always permissible to use wire with a higher temperature rating, but it must be sized in accordance with the temperature marking on the circuit breaker and *NEC* Table 310.16. If the width of the circuit breaker is 1-½ inches per pole or less, then this marking may be in any location except on the back (*NEC* 110.14(C)1).

8. HACR type – This marking indicates the circuit breaker is suitable for use with the group motor installations typically found in heating, air conditioning and refrigeration equipment. The *NEC* 2005 no longer has this marking requirement. The electrical industry determined that circuit breakers are considered suitable for use with such equipment without any further testing, therefore, the HACR marking is no longer required on air conditioning and refrigeration equipment or on circuit breakers for use in these applications. The requirement for this marking has also been removed from the UL 1995 product standard for HVAC equipment (see item 3 in photo 1).

Maximum wire size – Circuit breakers are typically marked with a wire range, however that marking is not mandatory. If the circuit breaker cannot accept the next larger wire size required for the ampere rating, then the maximum wire size must be marked in any location except the back (see item 5 in photo 3).

Separately shipped connectors –If connectors are not factory installed on a circuit breaker, then it must be marked with the proper connectors or terminal kits required in any location except the back (see item 8 in photo 3).

Ground-Fault Protection for People

The GFCI function, as part of a circuit breaker, provides ground-fault protection for people and has a number of unique marking and instruction requirements.

Test function – The GFCI has a test function that requires action upon installation and on a monthly basis. GFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate monthly testing.

Photo 4. 1Ø – 3Ø Marking
1. 1Ø – 3Ø marking – A 2-pole circuit breaker used to protect a 3-phase load on a corner-grounded delta system must be rated and marked for such an installation. Circuit breakers marked "1-phase – 3-phase" or "1Ø – 3Ø "are suitable for use on 3-phase corner-grounded delta or single-phase circuits (*NEC* 240.85).

Suitable for continuous operation at 100% of ampere rating only if used in a recognized enclosure 52 in. (1321 mm) by 20 in. (508 mm) x 7 1/2 in. (190 mm) deep or larger. Use only with 90° rated wire sized per the ampacity of 75° rated wire.

Figure 1. 100% Rated Marking
100 percent continuous rated – Circuit breakers are typically intended for use at not more than 80% of rated current where the load is considered continuous, or will continue for 3 hours or more (*NEC* 210.20). However, some circuit breakers are rated for continuous use at 100% of their current rating. These circuit breakers must be so marked in any location except on the back. Enclosure information such as a specific type or specific volume must also be marked. A requirement for the use of 90°C insulated wire sized to the 75°C column in *NEC* Table 310.16 and specific ventilation requirements may also be marked on the circuit breaker or equipment (*NEC* 210.20(A) and 215.3).

"Class A" marking – A "Class A" ground-fault device is intended to protect people. The Class A marking indicates that the trip threshold of the GFCI is between 4 mA and 6 mA. This marking may be in any location except the back.

Instructions – All GFCI circuit breakers must include instructions for the installer plus instructions on the use of the test function. A hangtag or self-adhesive label must also be provided, instructing the user to test the GFCI at least monthly. Inspectors should check to see that the tag or label has been properly installed.

Ground-Fault Protection for Equipment

Circuit breakers may also include a ground-fault protection for equipment (GFPE) function that, like GFCIs, has a number of unique marking and instruction requirements.

Test function – GFPE circuit breakers may have a test button or switch that may be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Trip level – GFPE circuit breakers must be marked with their trip threshold in milliamperes in a location accessible without removing trims or covers.

Instructions – All GFPE circuit breakers must include instructions for the installer.

Arc-Fault Protection

Circuit breakers may also include arc-fault protection (AFCI) that, like GFCIs, also has a number of unique marking and instruction requirements.

Device identification – AFCIs must also be identified appropriately. Branch/feeder or Combination type AFCIs must be so marked in a location visible when the trims or covers are removed. This is an important marking to note as we move into 2008, as *NEC*-2005 requires Combination AFCIs in bedrooms effective January 1, 2008 (*NEC* 210.12).

Test function – AFCI circuit breakers must have a test button or switch that must be labeled in a location accessible without removing trims or covers in order to facilitate testing.

Instructions – All AFCI circuit breakers must include instructions for the installer.

Circuit Breaker Markings Ensure a Safe Electrical Installation

So why are all of these markings on circuit breakers? Without them, it would be nearly impossible to install or inspect an installation for the appropriate performance ratings and fundamental electrical connections. When designing or completing an installation, key items to review are:

1. Are the voltage, continuous current, and interrupting ratings appropriate for the application?
2. Does the application require SWD or HID ratings?
3. Is the wire type and size appropriate for the circuit breaker?
4. Is the circuit breaker suitable for the equipment in which it is installed? Have other protective functions such as GFCI or AFCI been provided as required by the *NEC*?
5. Is the temperature rating of the circuit breaker suitable for the application?

The UL Marking Guide for Molded Case Circuit Breakers is a valuable resource to understand circuit breaker markings that may further explain these and other markings in detail. If you have questions about CB markings not answered here, consult the Marking Guide or the manufacturer to assist in an *NEC*-compliant installation.

Mr. Larsen, P. E., is the manager, Industry Standards for Circuit Protection for Square D Company / Schneider Electric. Mr. Larsen has served in various positions over the past 32 years. He is a member of several UL, IEEE and NEMA technical committees. His other memberships include NFPA, IAEI and CANENA.

Multiband and Silicon Integrated Antennas for Wireless Sensor Networks

by

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Ph.D. Dissertation

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Abstract

Wireless technologies have triggered an important revolution in the world of telecommunication networks and services. New systems, such as the latest generation mobile phones or wireless LANs, have appeared being enthusiastically accepted by people. In particular, applications related to remote sensor networks are growing rapidly (for temperature monitoring, the detection of free parking spaces, etc.).

This thesis is focused on antenna design for distributed wireless sensor networks. The remote sensors we are considering are connected to a central unit using a wireless link in the 5.8 GHz ISM band and the central unit transmits the collected information to the Internet via a WLAN connection.

In the central unit multiband antennas are necessary to cover several frequency bands with a single radiating element. Two solutions to obtain this multiband performance are proposed: fractal-based ground planes and resonator loaded monopoles. Several novel antennas that easily meet the system requirements are manufactured and measured.

For remote sensors small dimensions and low costs are desired. The use of silicon integrated antennas fabricated with standard CMOS technology is a good choice to achieve these requirements. First, the main problems of antenna integration are discussed and, then a possible solution to improve the antenna efficiency is presented together with a study of the effects of other elements integrated in the same chip with the antenna. The measurements of the manufactured prototypes demonstrate that antenna integration at 5.8 GHz is feasible, and that system ranges of some meters can be obtained even when the antenna is fabricated together with other elements integrated in the same chip.

The possibility of having some of the sensors implanted inside the body will be a reality in the near future, for this reason the particular case of implanted antennas is also analyzed. The last part of this thesis is devoted to study the body effects on antenna performance and evaluate the additional propagation losses introduced by the body tissues.

Agraïments

En aquesta vida tot té un inici i un final. Ara ha arribat un dels moments més esperats d'aquesta tesis, el moment d'escriure els agraïments, fet que indica que aquest treball ha arribat a la seva culminació. A més és l'ocasió perfecte per donar les gràcies a totes aquelles persones que m'han acompanyat durant aquests 4 anys.

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Pare, mare i avis, a vosaltres us deixo pel final. Vosaltres em vareu motivar a iniciar aquesta tesis i sempre heu estat presents en aquest llarg camí. Heu suportat el meu mal humor quan les coses no anaven com jo volia i sempre heu estat allà, animant-me a tirar endavant i a arribar al final d'aquest camí que vaig iniciar ja fa un bon temps. Moltes gràcies per tot, i recordeu que aquesta tesis va dedicada a vosaltres !!!!

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Radiation pattern measurements were carried out at the anechoic chamber of the Antenna Lab group of the Universitat Politecnica de Catalunya.

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Chapter 1

Introduction

1.1 Wireless communication systems: New trends and use of the ISM bands

1.1.1 New trends

Over recent years, wireless communication systems have spread widely and new communication services have been enthusiastically adopted by people around the world. Mobile communication systems have had a massive growth, firstly with voice services, and recently, with the latest generation of mobile data services that allow for high data transmission rates. Nowadays, it is quite common to find people reading their electronic mail, reading the last news in an online newspaper or updating their Facebook page in the train during their journey to the office.

The proliferation of computers and wireless communications has evolved into an era of wireless networking. Wireless Local Area Networks (WLANs) have replaced the traditional LANs in a huge number of applications because WLANs introduce benefits such as improved flexibility, mobility and ease of installation. Furthermore, the most modern commercial devices [1–4] are starting to use the latest generation mobile phone systems together with the WLAN systems to improve the user experience, providing access to the network inside and outside the office and taking advantage of the higher performance of WLANs when the user is inside.

Mobile phones and WLANs are probably the most popular wireless systems, but over recent years the use of other wireless devices has also increased. For instance remote sensor networks can be used to detect the free spaces in a parking lot and send this information to

the people that are looking for parking space [5–7]; or to precisely control the temperature at different points of a room or a building and adjust the air conditioner according to the sensors measurements [8].

In sensor networks a group of sensors are spatially distributed and, using a wireless link, they transmit the information collected to a central unit which can process it or retransmit it around the world over the Internet. In these applications sensor size and cost are key factors to ensure system viability. For this reason, the integration of the RF front-end together with the sensor elements can help obtain really small and cheap devices. Due to the increasing use of these systems, multiple commercial transceiver modules for these applications can be found [9]. However, the radiating element is usually an external antenna that increases the system size. Furthermore, costs are also increased by the need for a physical connection between the antenna and the integrated front-end. The integration of the whole system (including the antenna and the sensor) in a single chip can be advantageous and avoid the need for external connections.

Another field where the use of wireless devices is increasing is in medical applications, which can be understood as a specific case of sensor networks. The aim of these systems is to improve the life quality of patients, improving their mobility during treatment and the subsequent monitoring period. These systems can be also extremely useful for the treating of chronic diseases. As an example, wireless implantable glucose monitoring systems can improve the life and mobility of people with diabetes related problems. Moreover, these systems could also increase safety by alerting the emergency services when a problem is detected.

1.1.2 Use of the ISM bands

In recent years there has been a significant increase in the number of systems and devices that operate in the Industrial, Scientific and Medical (ISM) applications frequency bands. According to the Code of Federal Regulations (CFR) of the Federal Communications Commission (FCC) [10] and the International Telecommunications Union (ITU) regulations [11], the ISM bands are for operation of "equipment or appliances designed to generate and use locally RF energy for industrial, scientific, medical, domestic or similar purposes". Typical ISM applications are the production of physical, biological, or chemical effects such as heating, ionization of gases, mechanical vibrations, hair removal or acceleration of charged particles. Specific examples of the previous applications are domestic microwave ovens, jewelry cleaners for home use or ultrasonic humidifiers. Table 1.1 shows a summary of the frequency bands defined by the FCC for ISM

Table 1.1: *Frequency bands for ISM applications defined by the Federal Communications Commission (FCC) [12].* [1] *The use of these frequency bands for ISM applications shall be subject to special authorization.* [2] *Only in Region 2.*

Frequency Range	Center frequency
13553 - 13567 kHz	13560 kHz
26957 - 27283 kHz	27120 kHz
40.66 - 40.70 MHz	40.68 MHz
433.05 - 434.79 MHz	433.92 MHz[1]
902 - 928 MHz	915 MHz[2]
2400 - 2500 MHz	2450 MHz
5725 - 5875 MHz	5800 MHz
24 - 24.25 GHz	24.125 GHz
61 - 61.5 GHz	61.25 GHz[1]
122 - 123 GHz	122.5 GHz[1]
244 - 246 GHz	245 GHz[1]

applications, also known as ISM frequency bands.

The previous definition states that ISM bands were not originally specified for telecommunication services, but the FCC mentions that these bands can also be used for this kind of services [13]. However, the radio communication services operating within these frequency bands must accept harmful interference which may be caused by the other ISM applications. The FCC has several rules and regulations that must be satisfied by these communication systems to ensure their correct operation and avoid undesirable interferences to other systems [13].

Multiple license free communication systems that use some of the ISM bands have appeared in the last years. WLANs under the IEEE (Institute of Electrical and Electronic Engineers) 802.11b standard [14], one of the most widespread standards, use the 2.45 GHz ISM band for data transmission. Other examples can be found in Bluetooth systems [15], a wireless technology standard for exchanging data over short distances from fixed and mobile devices, or Zigbee systems [16]. Apart from these communication systems, different examples where the ISM bands are used in sensor networks can be also found in literature [17, 18].

Implantable devices could also operate in the ISM bands. Originally, the Medical Implantable Communications Services (MICS) 402 - 405 MHz frequency band was defined for these devices [19]. Some of the first examples of communication systems in implantable devices can be found in wireless links for pacemakers in order to monitor the pacemaker state [20]. Pacemakers are big devices, but in new applications, like implantable glucose monitoring systems, smaller devices are necessary. For that reason, the 2.45 GHz and

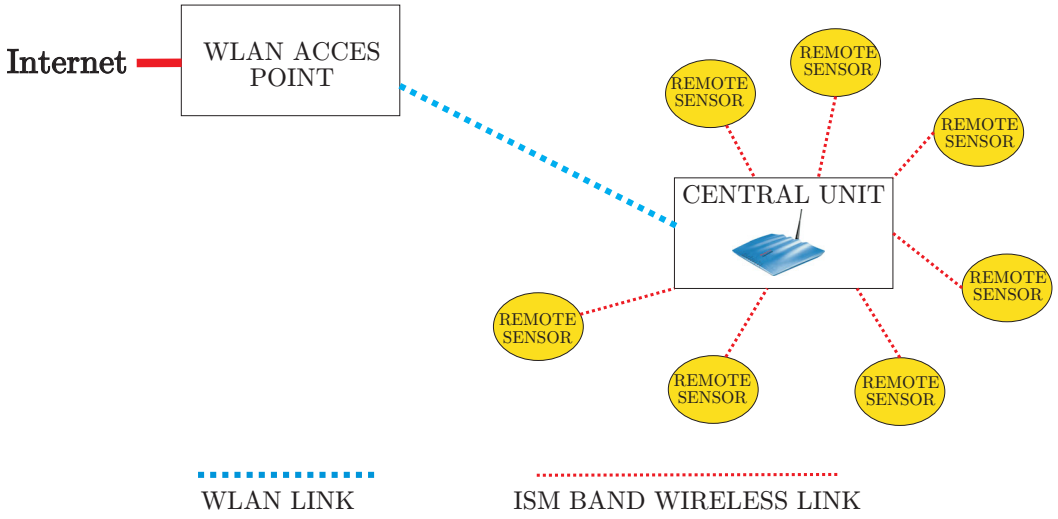


Figure 1.1: *Sensor network with a central unit that collects the sensors information and sends it to the Internet using a WLAN connection. The central unit and the remote sensors are connected using a wireless link working in the 2.45 GHz or 5.8 GHz ISM bands.*

5.8 GHz ISM bands could be considered to reduce the system dimensions.

1.2 Motivation and focus of this thesis

1.2.1 Thesis motivation and system description

This thesis is focused on the antenna design for a wireless remote sensor network with a central unit connected to the Internet through a WLAN access point (figure 1.1).

According to figure 1.1 all the sensors will be connected to a central unit and this unit will be connected to the Internet through a WLAN access point. Using the WLAN connection the central unit will be able to send the sensors information to anyplace around the world. In order to assure maximum coverage and compatibility with the latest standards, the central unit will use the 2.45 GHz and the 5 GHz bands defined by the 802.11 standards. With this dual band operation, the typical 802.11 a/b/g standards can be used, and furthermore the system will be also compatible with the most modern versions of the 802.11 standards, such as the 802.11n.

The remote sensors and the central unit will be connected through a wireless link using the ISM bands. To achieve small devices, the 5.8 GHz frequency band is advisable, but in some applications, such as implantable sensors, lower frequencies like the 2.45 GHz ISM

band can be desirable. Therefore the central unit must be able to simultaneously receive information from sensors operating in both bands.

The use of a central unit that collects the information instead of having the sensors directly connected to the Internet will allow for smaller and simpler sensors. Moreover, the distance between the sensors and the central unit will be smaller than the distance between the sensors and the WLAN access point, thus allowing for low power sensors that transmit tiny power levels, reducing power consumption and increasing the sensors battery life.

A particular case of the previous system is when one of the remote sensors is implanted inside the human body. An example of application with implanted sensors could be in a hospital, where they could be used to monitor the patients health status. Sensors would be connected to the central unit that could be a small receiver located on the patient bed or table. This central unit would be connected to the hospital network using the nearest WLAN access point, allowing the authorized doctors and nurses to access the patient information from any computer in the hospital. It is true that the central unit could be connected with a cable to the hospital network, but a WLAN connection has advantages, like the fact that the patient could be moved from one room to another (for instance for a special diagnostic operation) while keeping open the connection to the hospital network.

Multiple small variations could be introduced to the described system in order to modify its functionality. For instance a portable central unit with a Bluetooth connection could use the patient mobile phone to send the collected information to the Internet when away from a WLAN access point. Another option could be to use the new WiMAX (IEEE 802.16 [14]) systems to transmit the information to the Internet.

1.2.2 Focus of the thesis and objectives

The aim of this thesis is the design of antenna solutions to cover the needs of the sensor network presented in the previous section. The requirements of the central unit antenna and the sensors antenna are quite different; therefore both problems will be analyzed separately.

According to the above system description, the central unit must be able to use the 2.45 GHz and 5 GHz bands defined by the 802.11 standards, as well as the 5.8 GHz ISM band. One possible option could be to use a different antenna for each band; but this is not an optimal solution. Multiband radiators that can give service to several bands with a single radiating element are desirable. For this reason an important part of the work

presented in this thesis will be devoted to the design of multiband antennas to try to optimize performance while keeping to reduced dimensions. Among the possible solutions to achieve this multiband functionality, we will consider fractal geometries and resonator loaded structures. Our study will be focused on monopole antennas because, as can be seen in figure 1.1, the central unit will be similar to an ADSL router and monopole antennas are typically used in this kind of devices.

In remote sensors, small and cheap antennas are necessary. The integration of the RF front-end together with the sensors can be of interest to obtain small devices. However, the radiating element is usually an off-chip antenna thus increasing the system size and cost due to the need for a physical connection between the antenna and the RF front-end. A good choice to obtain really small systems is the use of system-on-chip solutions [21], where antenna and circuits are integrated together in the same chip. For this reason one chapter of this thesis will focus on the design of small antennas fabricated with standard CMOS silicon technologies. Furthermore, antenna performance degradation due to the presence of other components will be also analyzed.

The special case of a sensor implanted inside the human body will be considered. In implantable devices the human body effects on the antenna performance must be taken into account. Human tissues near the antenna will produce losses that will degrade the antenna performance; this being one of the most important limiting factors of the systems using implantable devices.

Therefore, the three main objectives of this thesis can be summarized as:

- Design of monopole multiband antennas for the central unit.
- Design of silicon integrated antennas for remote sensors.
- Study of the body effects when one of the sensors is implanted inside the human body.

Finally, we want to point out that the purpose of this thesis is not only to give a particular solution for a specific problem but to carry out a thorough study of the proposed topologies and its operating principles as well. Design guidelines will be given to assure that the topologies and results presented can also be applied in other systems with similar requirements.

1.3 Thesis organization

This thesis is divided into five chapters to gradually cover the aforementioned objectives.

Chapter 1, is the chapter that you are reading at this moment. It gives a brief introduction to the modern wireless communication systems and to the ISM bands. This chapter also presents our motivation and goals, showing the system architecture that has inspired and motivated this thesis. The results obtained during this thesis have permitted the author to publish and contribute to several papers which are detailed at the end of this chapter.

To satisfy the central unit requirements, high performance antennas with multiband behavior are necessary. For this reason, chapter 2 will focus on the design of multiband antennas covering the WLAN and ISM bands. Firstly a briefly introduction to multiband antennas will be given. Then, the chapter is divided in two sections. The first section analyzes multiband antennas inspired by fractal shapes. Some novel designs where fractals are used in the ground plane will be shown, and we will take advantage of the combination of the fractal shape in the ground plane and in the monopole element to improve antenna performance. In the second section, multiband behavior will be obtained using monopole structures loaded with resonators. Different novel topologies will be presented with a thorough analysis of their operating principles. Firstly the monopole will be loaded with a simple resonator and, next, it will be loaded with a dual resonator to introduce more operating bands. Finally, asymmetric structures will be also considered to improve multiband performance.

Chapter 3 will focus on the design of antennas for remote sensors. This chapter will analyze the design, fabrication and measurement of miniature integrated antennas fabricated with standard CMOS technologies. Firstly the performance offered by classical miniaturization technique when they are applied to antennas fabricated on a silicon substrate will be analyzed. Next, micromachining techniques will be suggested as a way to improve antenna performance and finally, the degradation introduced by other elements when they are fabricated together with the antenna will be studied.

In chapter 4 implantable antennas will be analyzed, focusing on the study of the propagation losses introduced by the human body. Furthermore, some of the results will be tested with measurements carried out using tissue simulating liquids.

Finally, chapter 5 includes the conclusions of this thesis and some guidelines for future work.

1.4 List of Author's Contributions

This section shows the contributions of the author to international journals and conferences during the period of development of this thesis. A distinction is made between those contributions related to the work of this thesis and those that are the result of the collaboration with other researchers but not directly related to the thesis content.

Journal Papers

- J. Gemio, G. Junkin, J. Parrón, R. Villarino, "Resonator-Loaded Dual-Band Monopole for Universal WLAN", *IEEE Antennas and Wireless Propagation Letters*, Vol. 8, 2009, pp. 736 - 739.

- J. Gemio, J. Parrón, J. Soler, "Dual Band Antenna with Fractal-Based Ground Plane for WLAN Applications", *IEEE Antennas and Wireless Propagation Letters*, Vol. 8, 2009, pp. 748 - 751.

- J. Gemio, J. Parrón, P. de Paco, G. Junkin, J. Marin, O. Menéndez, "A Split-Ring-Resonator Loaded Monopole for Triple Band Applications", *Journal of Electromagnetic Waves and Applications*, Vol. 24, 2010, pp. 241 - 250.

- J. Gemio, J. Parrón, and J. Soler, "Human Body Effects on Implantable Antennas for ISM Bands Applications: Models Comparison and Propagation Losses Study", *Progress in Electromagnetics Research*, Vol. 110, 2010, pp. 437 - 452.

- J. Gemio, J. Parrón, P. de Paco, J. Sacristan and A. Baldi, "Improving Silicon Integrated Antennas by Substrate Micromachining: A Study of Etching Patterns", *Progress in Electromagnetics Research*, Vol. 117, 2011, pp. 365 - 378.

Conference Papers

- J. Soler, J. Parrón, J. Gemio, "New Multiband Asymmetric Top-Loaded Mod-p Antenna for Terrestrial and Satellite Communications Services", *IEEE International Symposium on Antennas and Propagation*, San Diego, 2008.

- J. Gemio, J. Parrón, J. Soler, "Antena Dual con Plano de Masa Fractal para Aplicaciones WLAN", *XXIII Simposium Nacional URSI*, Madrid, 2008. **Best Paper Award: Premio de la Red Española de Antenas (RESA) a la mejor presentación de antenas**.

- J. Gemio, J. Parrón, J. Soler, "Multiband Antenna for WLAN Applications Using a Fractal-Based Ground Plane", *3rd European Conference on Antennas and Propagation*, Berlin, 2009.

- J. Gemio, J. Parrón, G. Junkin, "Monopolo Dual Cargado con Resonadores Magnéticos para Aplicaciones WLAN", *XXIV Simposium Nacional URSI*, Santander, 2009.

- J. Gemio, J. Parrón, J. Sacristan, F. Vila, F. Serra-Graells, A. Baldi, P. de Paco, "Miniaturization Techniques Applied to Standard CMOS Technology Integrated Antennas", *4th European Conference on Antennas and Propagation*, Barcelona, 2010.

- J. Gemio, J. Parrón, G. Junkin, "Resonator-Loaded Monopole Antenna with Triple Band Performance for WLAN/WiMAX Applications", *4th European Conference on Antennas and Propagation*, Barcelona, 2010.

- J. Gemio, J. Parrón, J. Sacristan, "Performance Degradation in Silicon Integrated Antennas due to Coils for Inductive Energy Coupling", *5th European Conference on Antennas and Propagation*, Rome, 2011.

Other journal papers

- O. Menéndez, P. de Paco, J. Gemio, J. Verdú, E. Corrales, "Methodology for Designing Microwave Acoustic Filters with Butterworth / Chebyshev Response", *International Journal of Microwave and Wireless Technologies*, Vol. 1, 2009 , pp. 11 - 18.

- O. Menéndez, P. de Paco, J. Gemio, "General Equations for the Design and Evaluation of a Stack of Two Bulk Acoustic Wave Resonators", *Microwave and Optical Technology Letters*, Vol. 51, 2009, pp. 752 - 755.

- M. Ramírez, J. Parrón, J. M. González-Arbesú, J. Gemio, "Concentric Annular-Ring Microstrip Antenna with Circular Polarization", *IEEE Antennas and Wireless Propagation Letters*, Vol. 10, 2011, pp. 517 - 519.

Other conference papers

- J. Gemio, O. Menéndez, P. de Paco, "Performance Evaluation and Design of Stacked Crystal Filter", *European Microwave Week 2008*, Amsterdam, 2008.

- A. Arriola, G. Sasiain, J. I. Sancho, J. Parrón, J. Gemio, R. Villarino, "Compact High Impedance Surface Based on Interdigital Capacitors", *4th European Conference on Antennas and Propagation*, Barcelona, 2010.

Chapter 2

Multiband antennas for the central unit

This chapter will focus on the design of multiband antennas for the central unit of the remote sensor network. As it has been said in the introduction, multiband performance is necessary to serve the ISM and WLAN frequency bands with a single radiating element.

Firstly a brief introduction to the concept of multiband antennas will be given and previous works related to multiband radiators will be shown. Next, two options to solve our problem will be analyzed: fractal shapes and resonator loaded structures. Some novel antenna designs will be presented for each option. Although this chapter is focused on monopole configurations, the same ideas could also be applied in the design of other kinds of antennas, such as dipoles or PIFA antennas.

2.1 Introduction to multiband antennas

Some of the classic antennas that can be found in literature, like straight dipoles or horns, have a behavior highly dependent on frequency [22–24], which means that they only work properly at the design frequency. However, it would be desirable to have antennas that could work simultaneously at multiple frequencies. For this reason multifrequency antennas are required.

There are two types of multifrequency antennas: multiband and broadband antennas. While a multiband characteristic is related to a discrete set of frequency bands where the antenna parameters are similar [25], broadband behavior is associated with an antenna whose parameters are similar within one large frequency band [22].

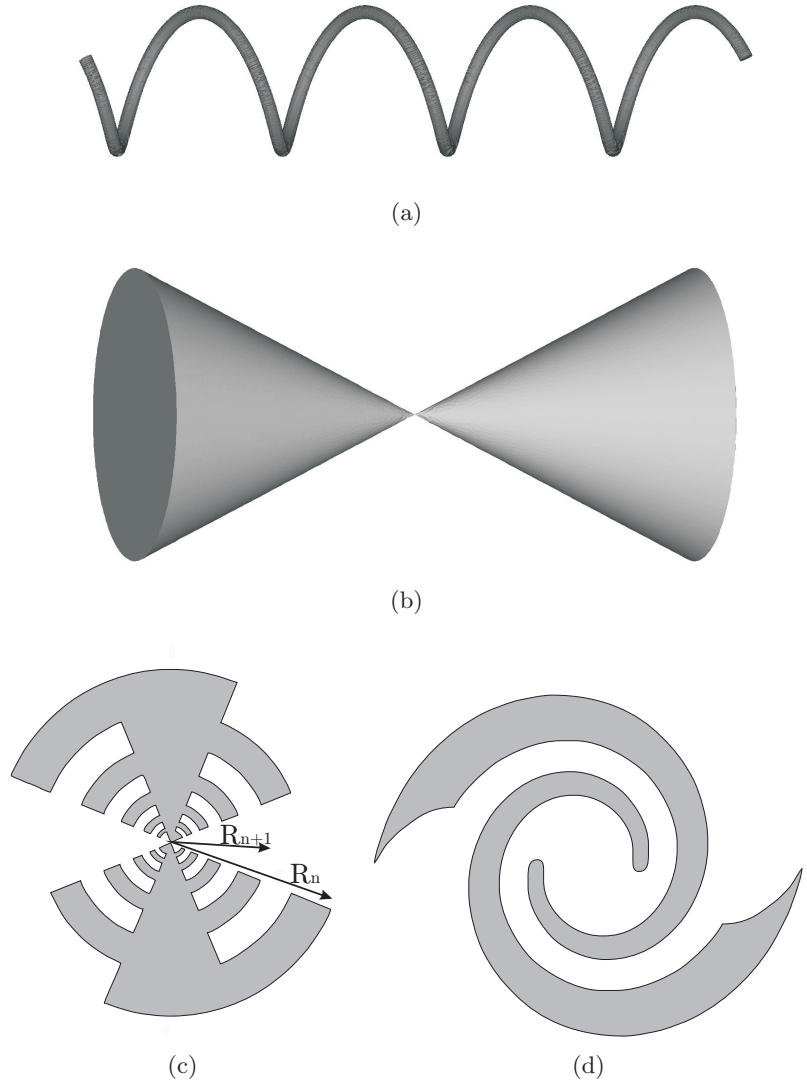


Figure 2.1: *Classic broadband and multiband antennas. (a) Helix antenna. (b) Biconical antenna. (c) Log-periodic antenna. (d) Spiral antenna.*

Different basic topologies have been used to obtain multiband or broadband behavior. Some well known antennas with broadband behavior are the helix (figure 2.1(a)) and the biconical antennas (figure 2.1(b)). To simplify the fabrication process, some geometrical approximation to the biconical antenna, like the triangular monopole or the bow tie dipole, can also be used [22].

The self scalable principle of antennas states that the input and radiation parameters of an antenna are invariant to a change of its physical dimensions if a similar change is also made in the operating frequency [22]. Based on this idea, DuHamel [26] introduced the concept of log-periodic antennas. An example of a real implementation can be seen in

figure 2.1(c), where the geometric ratio (τ) is defined as

$$\tau = \frac{R_n}{R_{n+1}} = \frac{f_{m+1}}{f_m} \tag{2.1}$$

being f_m and f_{m+1} the frequency of two consecutive operating bands of the antenna. In this kind of antennas, the geometric ratio (τ) observed in the geometry also defines the antenna log-periodic behavior in frequency.

Based on the same idea, in 1957 Rumsey [27] introduced the frequency independent antennas. He said that the behavior of an antenna that could be fully described by angles would be frequency independent. This is a theoretical concept, nevertheless there are some real implementations, like the spiral antenna of figure 2.1(d), that come close to this idea.

These are just a few examples of classical topologies used for broadband and multiband antennas. They were presented half century ago, but they are still in use, and some of them have been the base for the development of modern multiband antennas, such as the fractal antennas that will be analyzed later in this chapter.

Over the past few years an important effort has been put into the design of new multiband antennas for multiple system coverage in order to satisfy the needs of the market, which demands high performance small radiators with multiband behavior. Multiple examples of multiband antennas can be found, with designs that range from those based on monopoles with stub tuning [28, 29], sleeve dipoles [30] or multi-branch [31], to those based on dielectric resonators [32, 33]. Recently, the use of PIFA antennas [34–36] for mobile and portable terminals has been increasing due to their compact size and ease of integration inside the mobile terminal. Figure 2.2 shows some examples of the aforementioned solutions.

Top-loading techniques are another interesting option to achieve multiband behavior. These techniques have been typically used to reduce antenna dimensions [22]. For instance the introduction of a top capacitive load in a monopole antenna increases its electrical length, reducing the operating frequency without increasing the monopole height. However, top-loading techniques can also be applied to some antenna geometries, such as a triangular monopole or a Sierpinski monopole, in order to obtain multiband radiators [37]. From figure 2.3 it is seen that when loads are introduced to the antenna geometry, new current paths are created thus adding new resonances to the antenna response.

It is advisable to use these antennas at even resonances because relative changes in the input resistance are smaller than at odd resonances, but these resistance values differ from 50 Ω and hence the matching levels are poor. As was suggested in [37] the input resistance can be adjusted by modifying the antenna geometry. Figure 2.4 shows a modified version

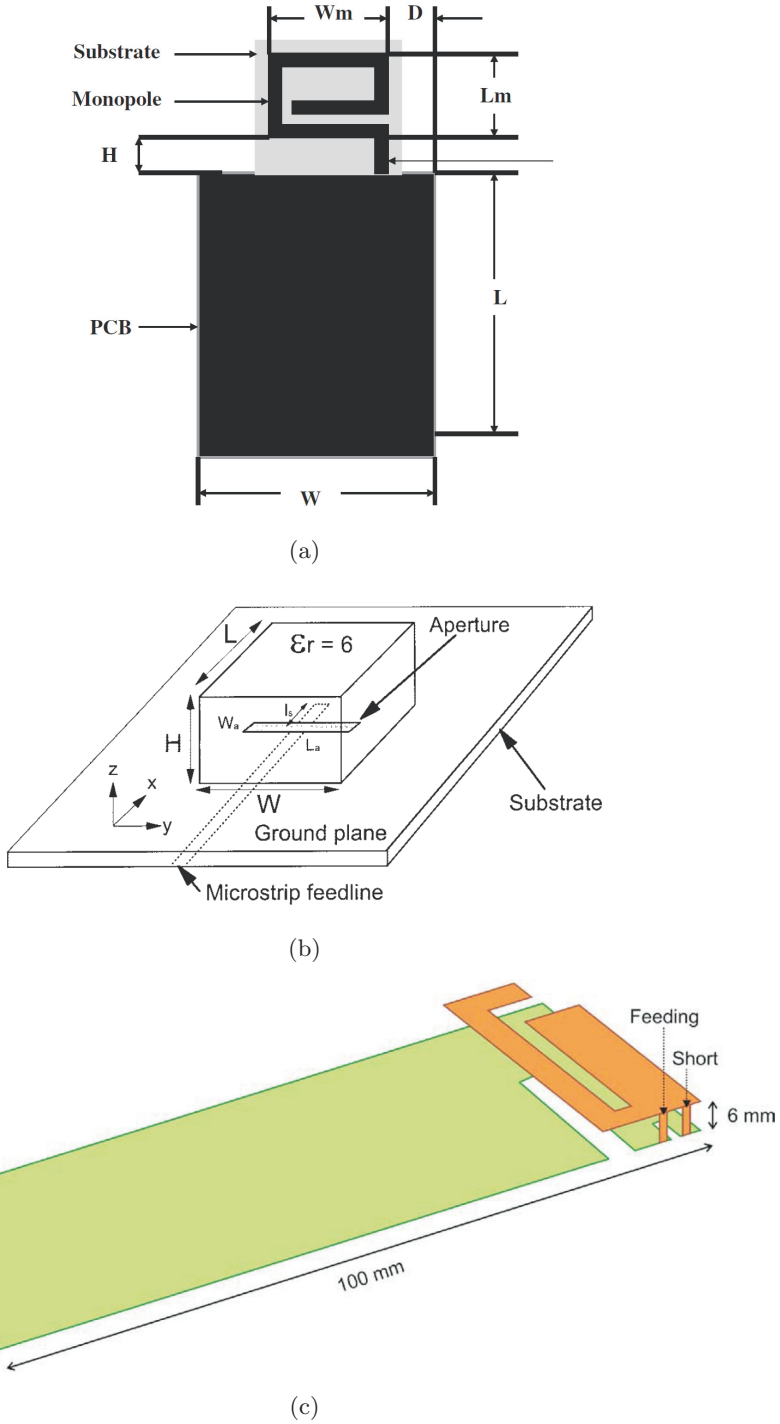


Figure 2.2: *Multiband antennas presented in recent years. (a) Monopole with stub tuning [28]. (b) Antenna based on dielectric resonators [32]. (c) PIFA Antenna [35].*

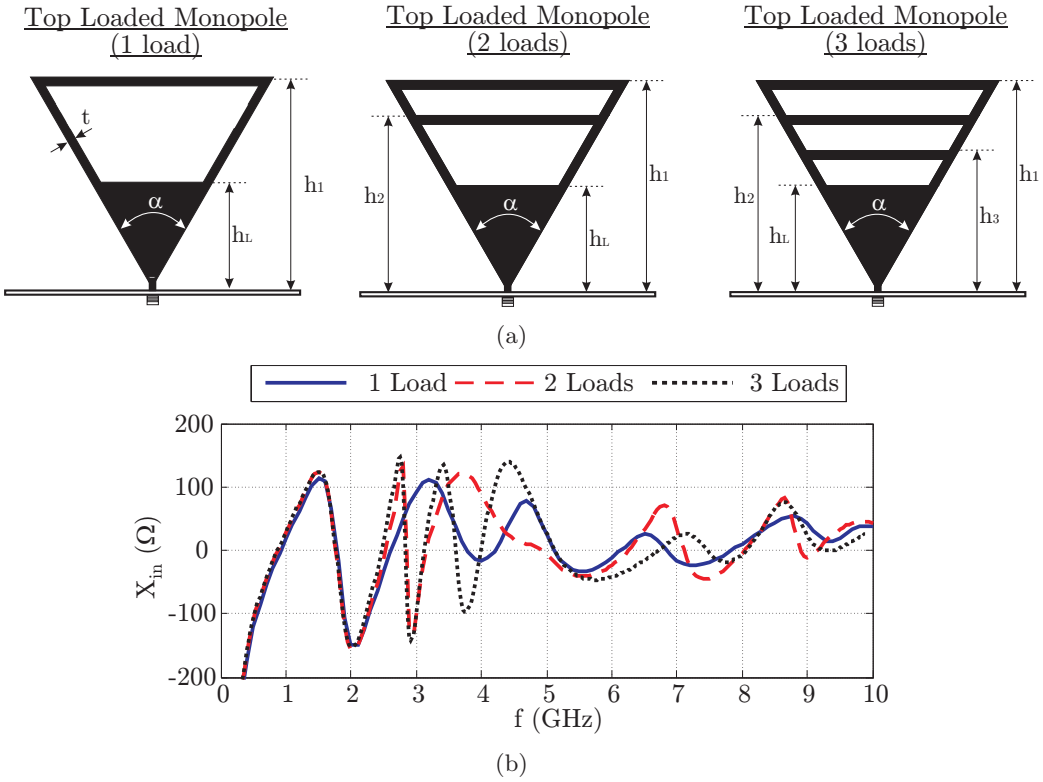


Figure 2.3: *Top-loaded triangular monopole with different load numbers. (a) Antenna geometry. (b) Antenna input reactance.*

of the triangular monopole antenna where the width w has been increased to reduce input resistance at even resonances and improve matching levels. Input reflection coefficients better than -10 dB are achieved in the four antenna matching bands.

It is important to point out that, in the framework of this thesis, similar input parameters in all the bands will not be the only design requirement for our multiband antennas. Similar radiation patterns will also be necessary.

2.2 Multiband antennas inspired by fractal geometries

In this section, it will be shown that fractal shapes can improve multiband antenna performance when they are used either in the monopole element or in the ground plane generation process. Several antenna designs suitable for the central unit of the wireless sensor network will be presented.

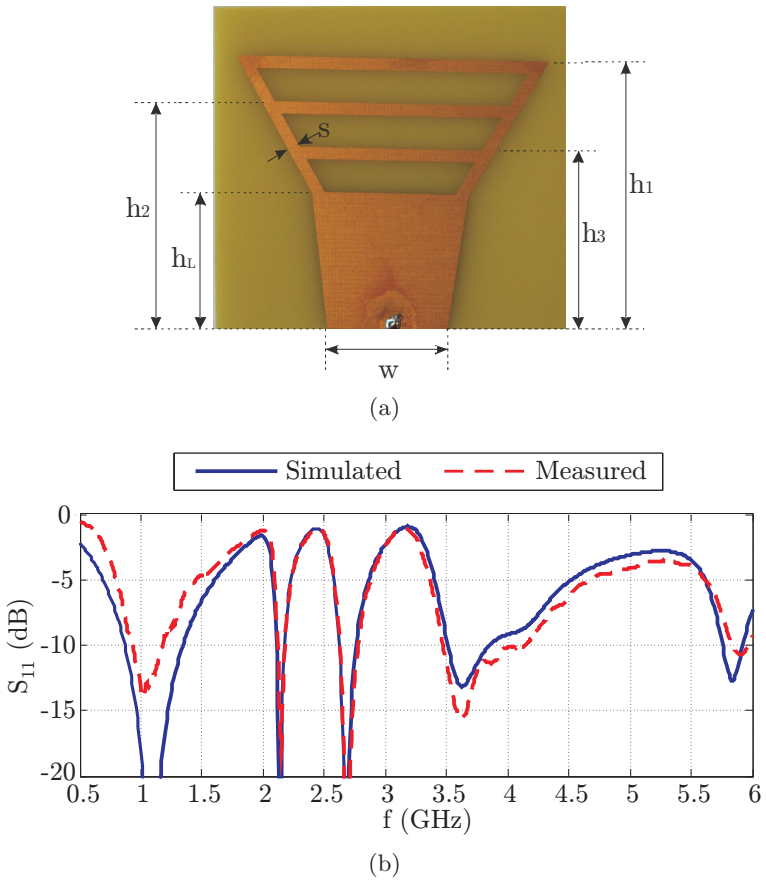


Figure 2.4: *Modified top-loaded triangular monopole. The width w has been increased to improve the matching levels. (a) Fabricated prototype. h_1 = 50 mm, h_2 = 41.7 mm, h_3 = 33.3 mm, h_l = 25 mm, s = 2 mm, w = 25 mm. (b) Measured and simulated input reflection coefficient.*

2.2.1 Introduction to the use of fractal shapes in antenna design

The definition of fractal is not a clear concept. The most important properties that characterize a fractal are its fractal dimension and its self similarity property [38–41]. Mandelbrot gave the following definition: "A fractal is by definition a set for which the Hausdorff dimension strictly exceeds the topological dimension", which he later replaced with: "A fractal is a shape made of parts similar to the whole in some way" [39–41]. In [39] it is said that the condition that must be accomplished to define an object as a fractal is that its fractal dimension exceeds its topological dimension.

One of the main ideas behind the fractal concept is that most fractal objects have self similarity at different scale levels, which means that a fractal object is composed of small but similar copies of itself. An example of a self similar fractal shape in nature is the

cauliflower (figure 2.5). If it is analyzed at different scale levels, the same shape is observed. In nature, other examples of fractal shapes can be found in trees, leaves, snowflakes or the coastline, to mention just a few examples.

Fractal shapes can be generated with an iterative process where each iteration is created using multiple rotated and/or scaled copies of the previous iteration. As can be seen in figure 2.6(a), the Sierpinski gasket generation process starts with a solid triangle. In each iteration, the new geometry is generated using three copies scaled by a factor of 2 of the previous iteration. Figure 2.6(a) shows the evolution of the geometry when the number of iterations is increased. The fractal shape is the geometry obtained after an infinite number of iterations. The generation process of other fractals, like the Sierpinski carpet or the Koch curve, is also shown in figure 2.6.

Strictly speaking, fractal shapes can not be physically built since they are a mathematical abstraction which includes an infinite number of elements. However, some real structures inspired by these ideal fractals are typically used. These real implementations are usually known as truncated fractals, prefractals or fractals with a finite number of iterations [39].

The use of fractal shapes is widespread in antenna design. One interesting property of some fractals, like the Koch or the Hilbert curves, is the space filling property which says that they can have an infinite length and be fitted in a finite area. This property is useful to reduce the antenna dimensions [42, 43], because when the number of fractal iterations is increased, the total antenna length is also increased, reducing the resonance frequency

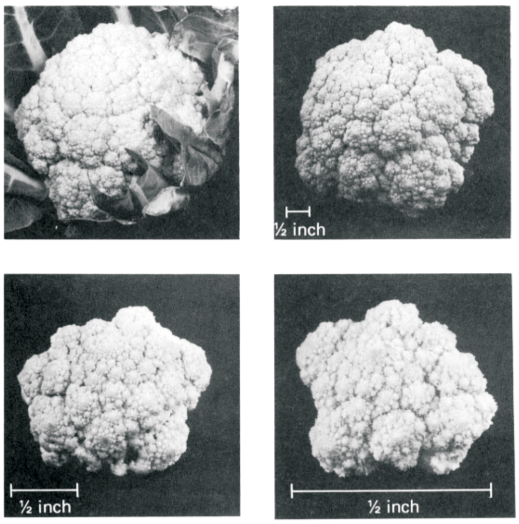


Figure 2.5: *An example of fractal shape in nature: Cauliflower [38].*

without a significant change in the area.

The self similarity property of some fractals makes these structures very suitable for multiband antennas. One of the first reported examples of a multiband fractal-based antenna was the Sierpinski antenna shown in [44, 45], where the Sierpinski gasket was used to obtain a multiband monopole. Since these initial references, further works have been presented in order to improve the Sierpinski monopole performance and have more control in the allocation of the operating bands [46–53].

Most of the aforementioned examples are based on dipole or monopole configurations; however similar ideas can be applied to other types of antenna. For instance in [54–56] the Sierpinski gasket (or a modified version of it) is used in the design of multiband patch antennas.

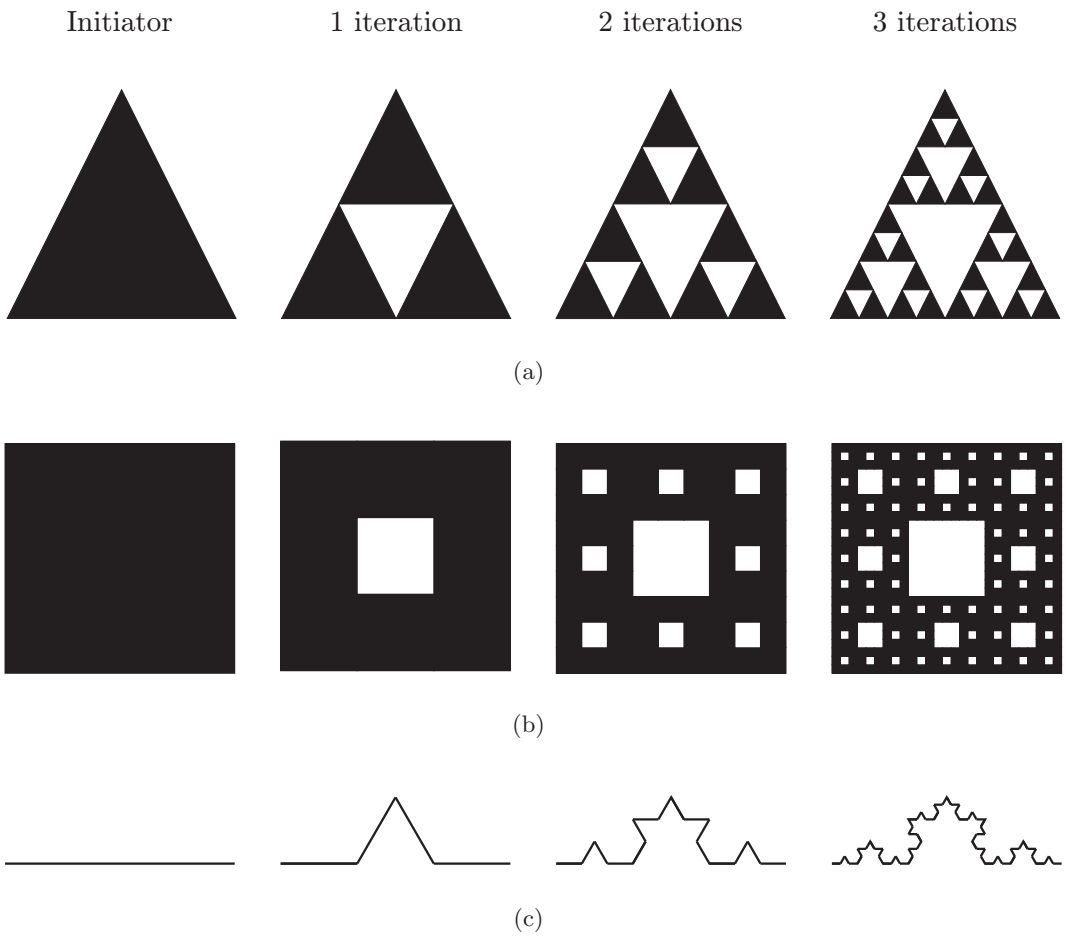


Figure 2.6: *Generation of different fractal shapes using an iterative process. (a) Sierpinski gasket. (b) Sierpinski carpet. (c) Koch curve.*

2.2.2 The Sierpinski monopole

In this section the performance of the Sierpinski monopole will be further analyzed because some of the new antennas presented in this thesis will be based on this structure. In the Sierpinski monopole the number of fractal iterations is directly related to the antenna multiband behavior. When additional fractal iterations are applied to the geometry, new resonances are introduced to its response.

According to figure 2.7 when the number of iterations is equal to two, three pairs of resonances can be clearly seen at low frequencies (0 - 6 GHz). When the number of iterations is increased to three, a new pair of resonances appears at higher frequencies (8 - 12 GHz). This tendency goes on as more fractal iterations are included in the antenna geometry. Good matching levels are obtained at the odd resonances when the reference impedance is 50 Ω, therefore the position of the operating bands will be mainly defined by the odd resonances of the structure. Moreover, if radiation patterns at the odd resonances are analyzed, acceptable similarity between them is obtained, and hence the Sierpinski monopole can be considered a multiband radiator [45].

The Sierpinski monopole multiband behavior is related to the self similarity property at different scale levels of this geometry and it can be easily understood analyzing the current distributions of the antenna (figure 2.8). When the frequency increases, currents are concentrated in a smaller part of the antenna, thus the antenna active area (area where currents are important) scales with frequency. If we zoom in on the active area, the shape of the current distributions at the different resonances is the same but scaled. This similarity in the current distributions allows for the antenna multiband behavior, offering similar input and radiation parameters in multiple operating bands. In the particular case of figure 2.8 the antenna will present a frequency log-periodic behavior, with a period equal to 2, due to the fact that the geometry has been created with a scale factor of 2. It is important to point out that the first resonance behavior will be slightly different due to the truncation effects related to the finite dimensions of the real antenna [44, 45].

The antenna of figure 2.7 is the basic Sierpinski monopole, however multiple perturbations can be applied to the original geometry to change and adjust the position of the bands, the periodicity among them and the radiation patterns shape. In the Sierpinski monopole the resonance frequencies are associated to the monopole edge length rather than to its height [48]. For a fixed height, when the flare angle α is increased, the edge length is also increased and the resonances will be shifted to lower frequencies. Another important factor is that the flare angle also modifies the radiation pattern shape. According to figure 2.9 when the flare angle $\alpha = 90^{o}$ the monopole offers a bidirectional radiation

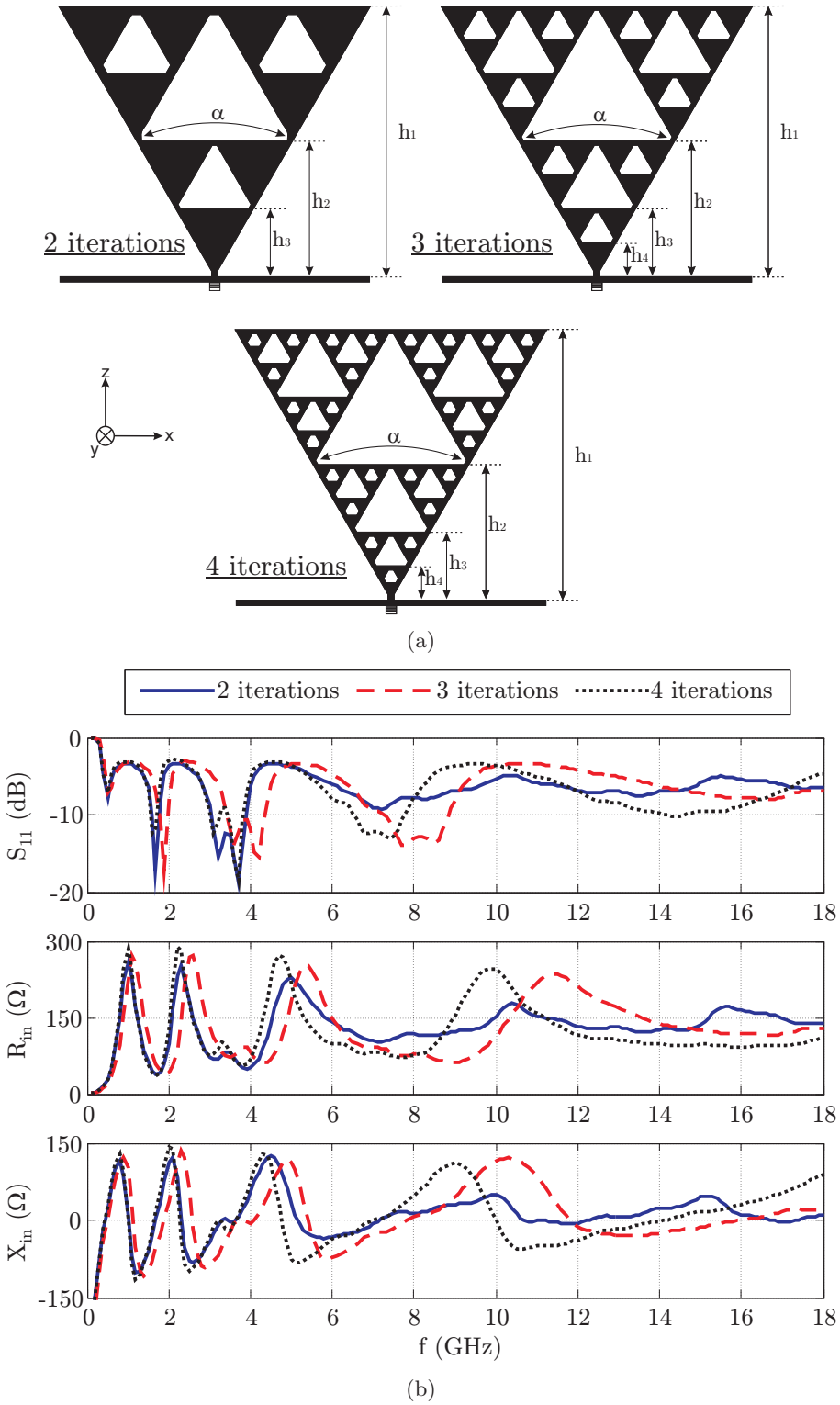


Figure 2.7: *Sierpinski monopole antenna with different numbers of iterations.* $h_1 = 100$ *mm,* $\alpha = 60^o$ *and* $h_{n+1} = h_n/2$.

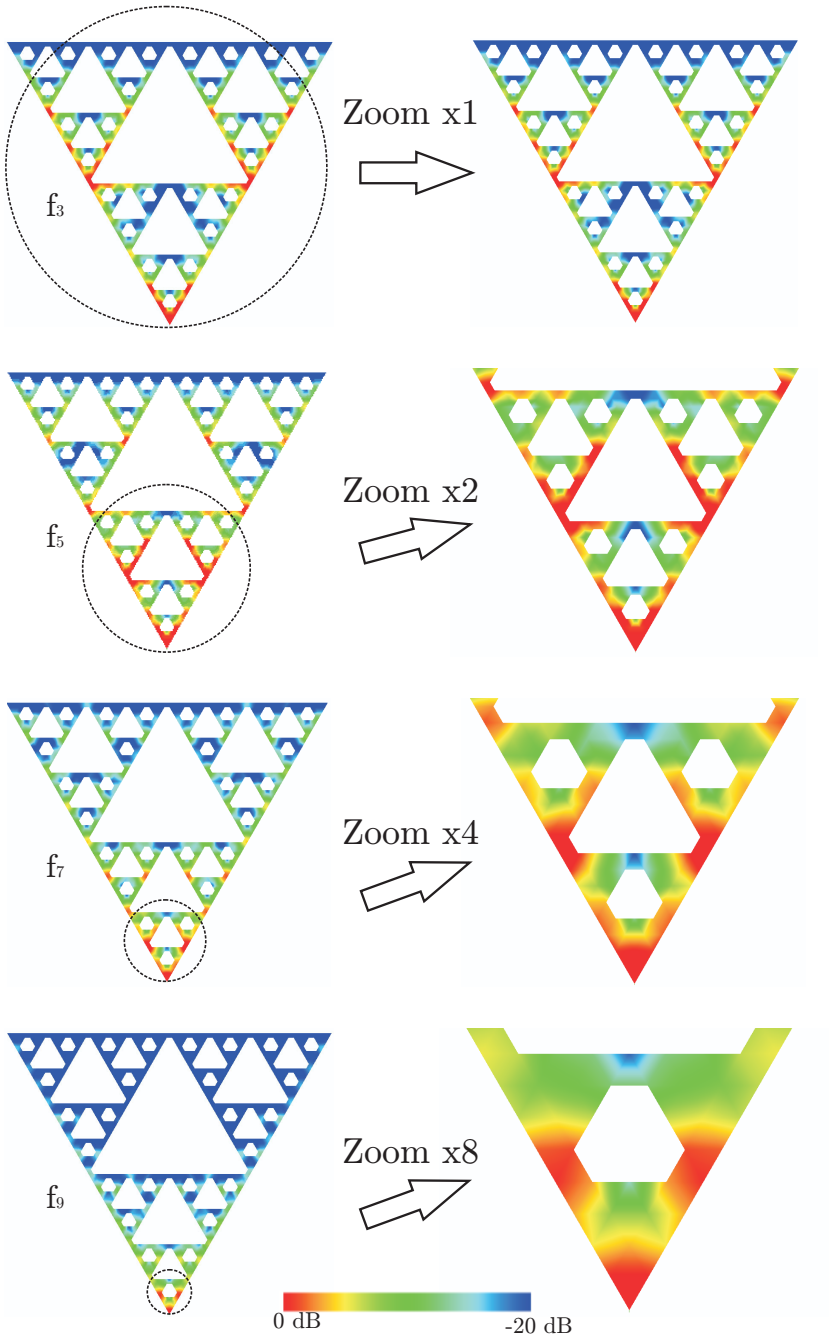


Figure 2.8: *Normalized current distributions at the odd resonances of a Sierpinski monopole with 4 fractal iterations.*

patterns, whereas if this angle is increased to $\alpha = 110^{o}$ radiation patterns become quad-directional. Therefore, depending on the application, the flare angle can be changed to adapt the radiation patterns to the specific application requirements.

Flare angle and antenna height can control, the radiation patterns and the operating bands position while keeping the same frequency spacing. The frequency spacing can be adjusted by changing the scale factor of the fractal geometry (figure 2.10(a)) [46]. Another option is the structure known as Mod-p (figure 2.10(b)), where multiple bands and sub-bands can be defined [49, 50].

As is shown in figure 2.11, the Sierpinski monopole can be combined with the aforementioned top-loading techniques with the aim of improving its multiband behavior [57]. Top-loading resonances are added at low frequencies and can be used to adjust the frequency spacing between the two first bands which does not agree with the scale factor because of the truncation effect of the fractal geometry. At high frequencies the top-loading effects will be less important because the antenna currents self scale with frequency approaching the antenna feeding point.

Furthermore asymmetric structures can increase the design flexibility, due to the fact that the whole structure resonances will combine resonances of both halves. As an example, figure 2.11 shows an asymmetric antenna based on two halves of top-loaded Sierpinski monopoles of different heights [57]. The asymmetric antenna reactance and the reactance of the isolated halves are plotted in figure 2.12. At low frequencies the input impedance of the asymmetric structure is equal to the parallel of the impedance of both isolated halves. For this reason the odd resonances of the asymmetric structure approximately coincide with the odd resonances of the two separated halves.

The relative variations in the real part of the impedance are generally smaller in even resonances than in odd resonances. Therefore to obtain multiband behavior the use of even resonances is advisable because it will allow us to obtain good matching levels simultaneously in the different bands if the average value of the input resistance at these resonances is used as reference impedance. Good antenna matching is achieved when the reference impedance is 170 Ω (figure 2.13).

According to figure 2.14, radiation patterns at even resonances are not equal in all the bands, but it could be useful because two different patterns are obtained. At the first three resonances the typical monopole-like pattern is obtained. This pattern could be suitable for terrestrial communications. At the fourth and fifth even resonances, the radiation levels for higher elevation angles increase, and this is useful for satellite communications. These results open the possibility of using the symmetric antenna as a solution for terrestrial

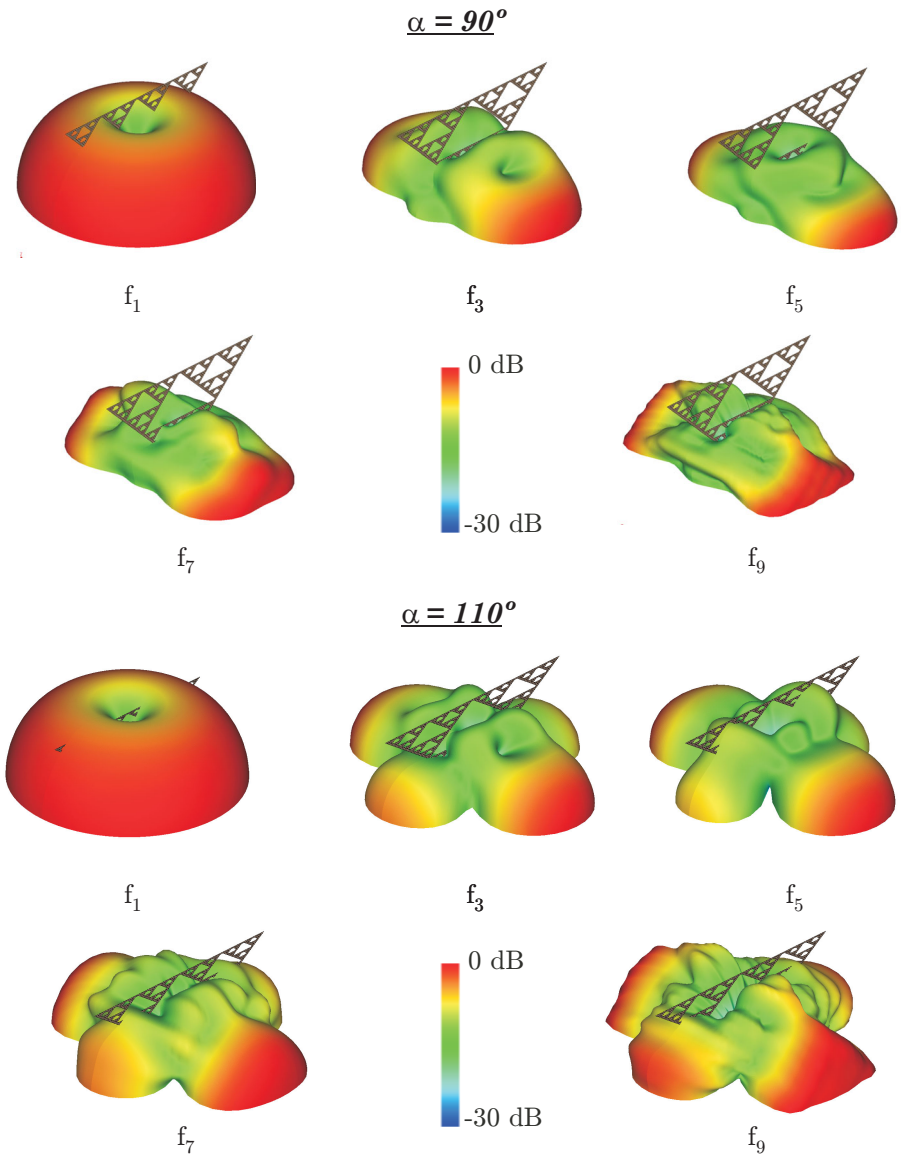


Figure 2.9: *Radiation patterns at the odd resonances of the Sierpinski monopole with different flare angles* α.

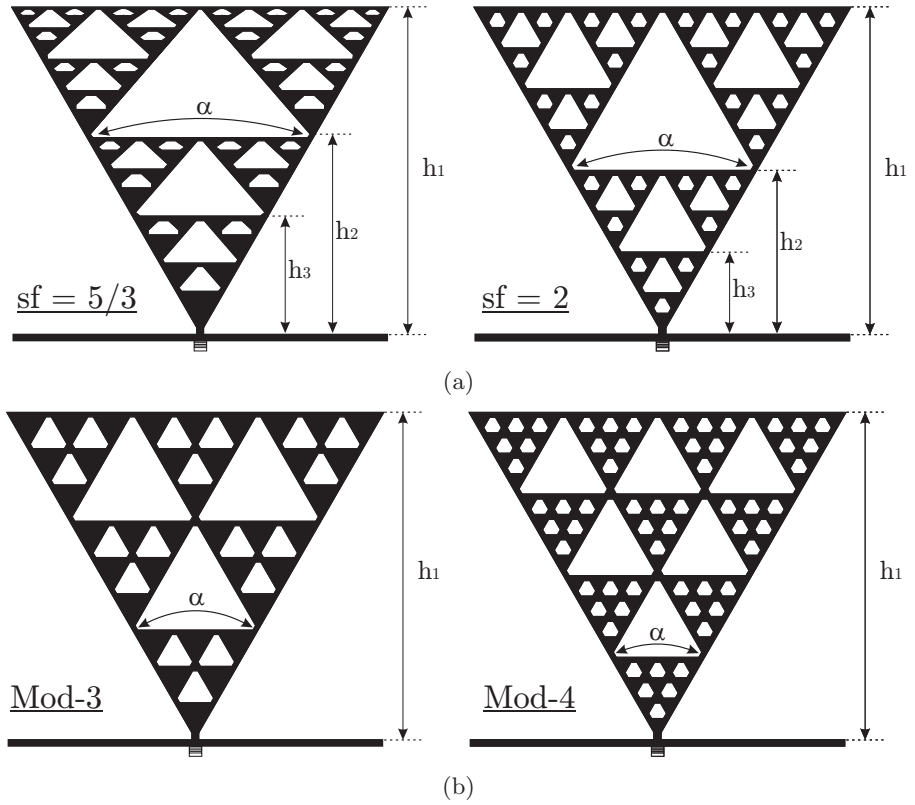


Figure 2.10: *Sierpinski monopole perturbations to control the spacing between resonances: (a) The scale factor $sf = h_n/h_{n+1}$ can be changed (b) or different Mod-p structures can be used.*

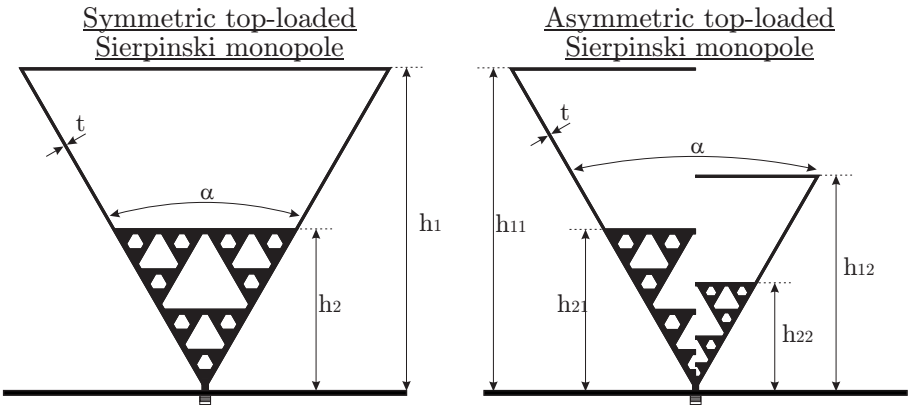


Figure 2.11: *Top-loaded Sierpinski monopole. $h_{11} = 300$ mm, $h_{21} = 150$ mm, $h_{12} = 200$ mm, $h_{22} = 100$ mm, $\alpha = 60^o$ and $t = 2$ mm.*

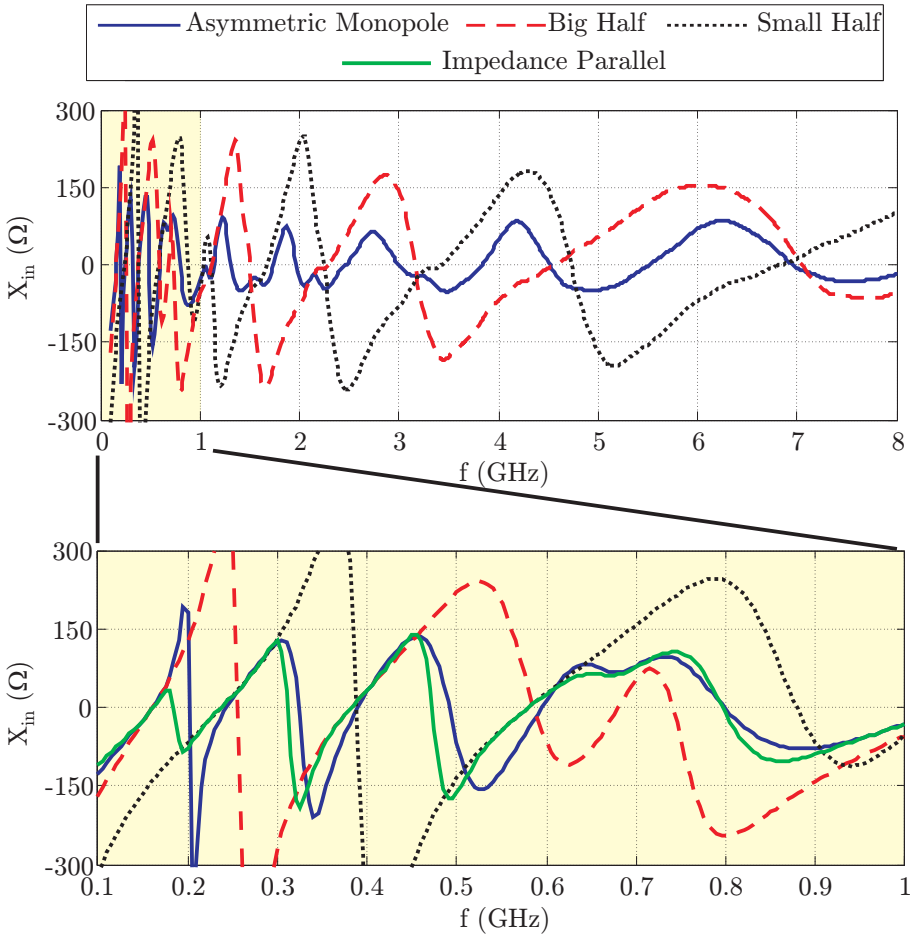


Figure 2.12: *Input reactance of an asymmetric top-loaded Sierpinski monopole (figure 2.11). The input reactance of the two isolated halves and the reactance of the parallel of the two halves impedance are also plotted.*

and satellite services with a single radiating element.

In the sensor network these radiation patterns could also be of interest. For instance in an outdoor sensor network where there is no WLAN or mobile phone system coverage, the low frequency bands could be used to collect information from the sensors, and this information could be transmitted to the Internet using a satellite communication system operating in high frequency bands. In this scenario, the satellite communication system would replace the WLAN connection of the original system. Another possible application could be in indoor scenarios where different frequency bands could be used to receive information from sensors at different elevation angles. However, in order to use this antenna in the sensor network, the antenna geometry should be adjusted to correctly allocate the operating bands according to the system requirements.

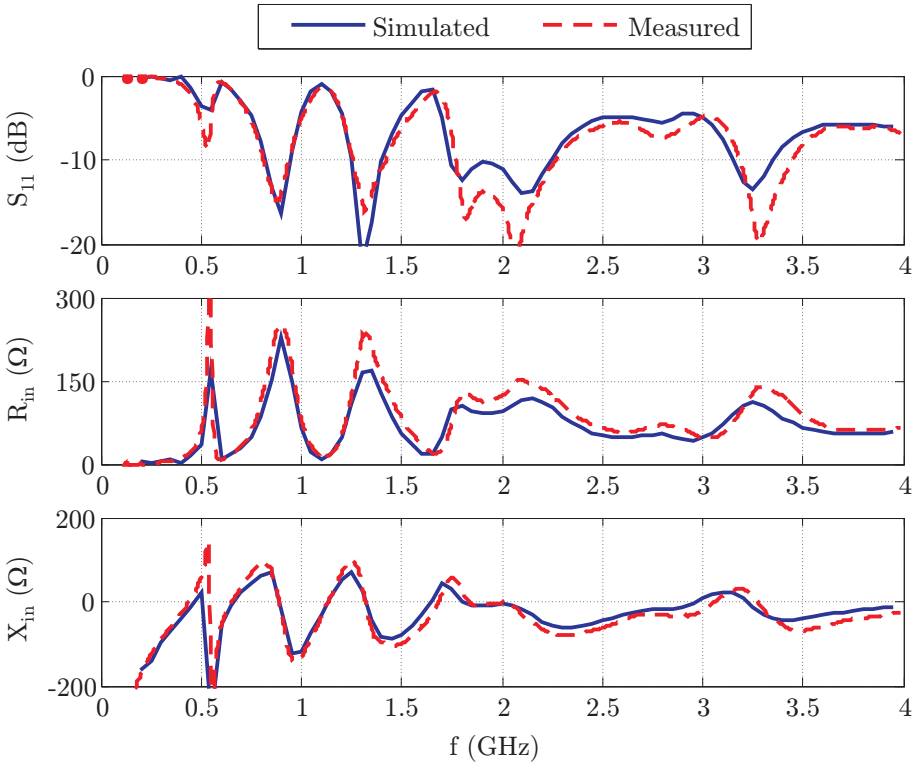


Figure 2.13: *Input parameters of an asymmetric top-loaded Sierpinski monopole (scaled prototype with h_{11} = 90 mm). The reference impedance for the S_{11} parameter calculation is 170 Ω instead of 50 Ω.*

2.2.3 Use of the Sierpinski gasket in the ground plane

In the previous section, and in most of the previous works presented in literature, the Sierpinski gasket is used in dipoles or monopoles with solid ground planes. In this section the fractal shape will be used in a different way, as part of the antenna ground plane.

If not properly taken into consideration, a ground plane with finite dimensions can introduce undesirable effects and degrade antenna performance. Nevertheless, it will be shown that we can make good use of the finite dimensions ground planes effects to improve antenna performance. In particular we will use a finite ground plane inspired by the Sierpinski gasket to improve the multiband behavior of the whole radiating structure.

The ground plane resonances will be used to reinforce the monopole resonances, improve the matching levels and bandwidths, and even introduce new operating bands. Several examples that use the ground plane effects to improve antenna performance can be found in literature [35, 58–63], but in these references the ground plane is not based on fractal shapes, which is one of the main novelties presented in this thesis.

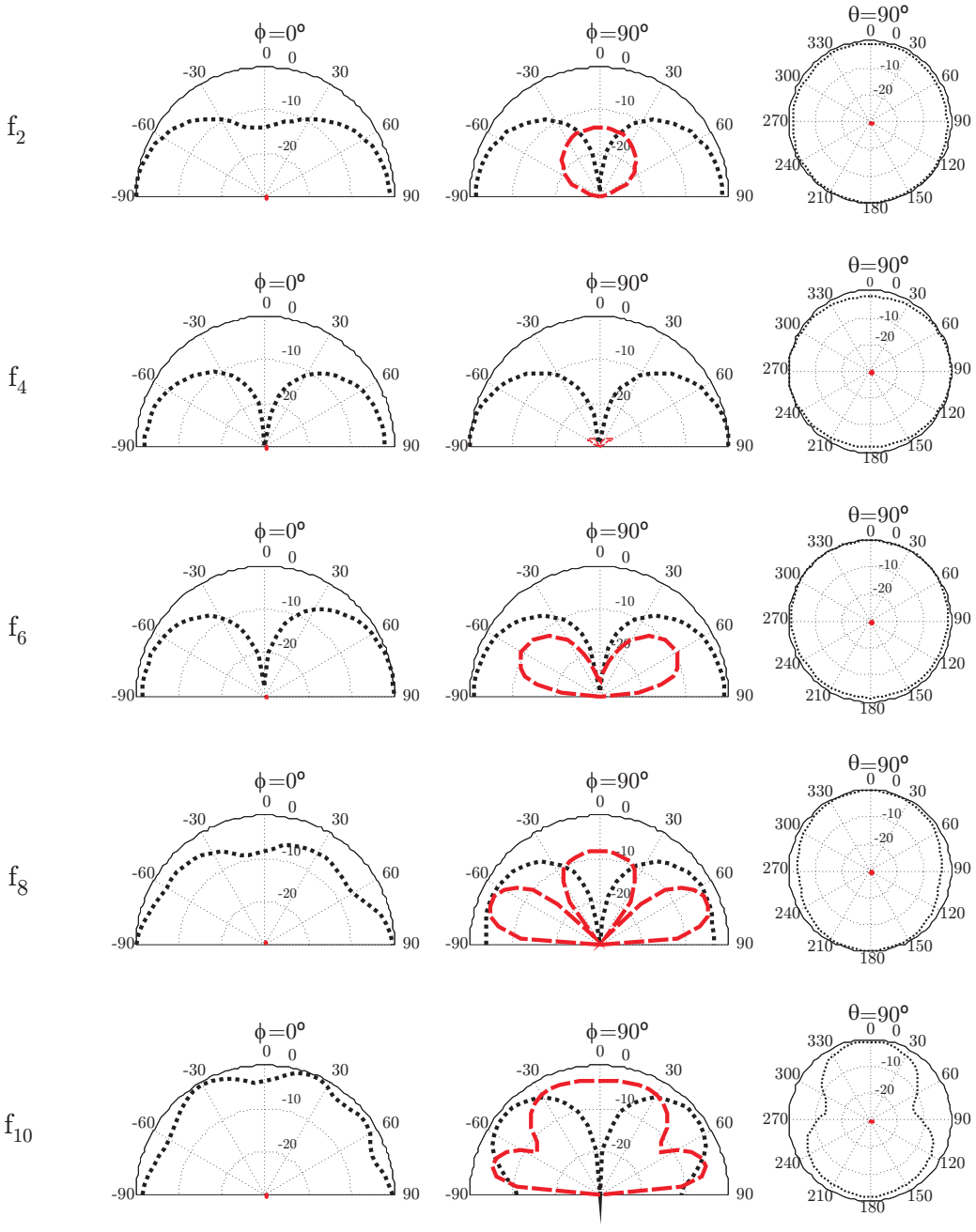


Figure 2.14: *Radiation patterns at the even resonances of an asymmetric top-loaded Sierpinski monopole (figure 2.11). Black dotted lines are used for the copolar component; red dashed lines are used for the crosspolar component.*

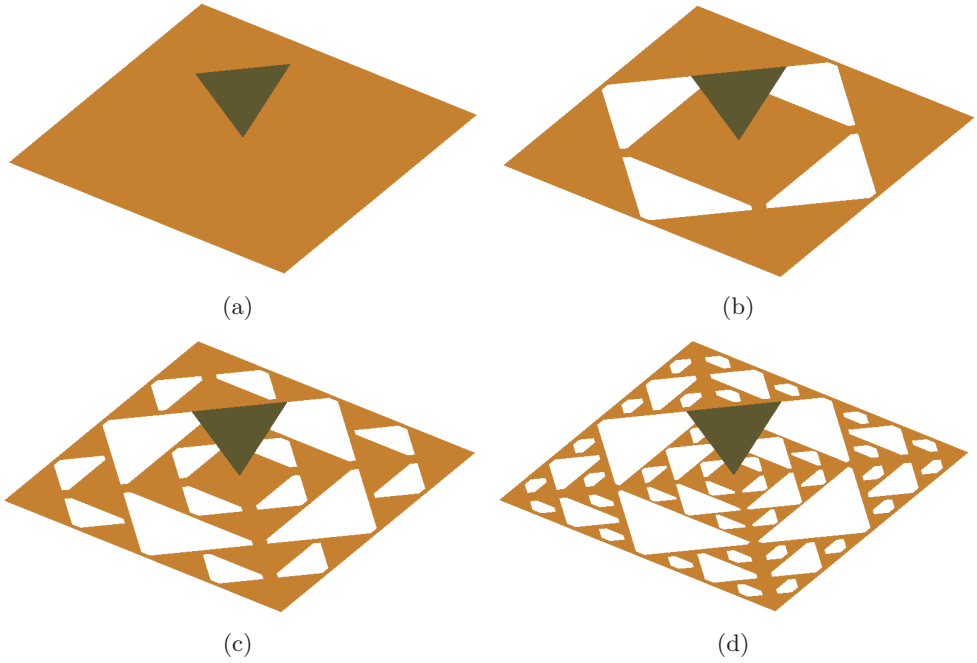


Figure 2.15: *Triangular monopole antenna on a finite dimensions ground plane.* (a) *Solid ground plane.* (b) *Fractal-based ground plane after 1 iteration.* (c) *Fractal-based ground plane after 2 iterations.* (d) *Fractal-based ground plane after 3 iterations. The triangular monopole height is 50 mm and the ground plane dimensions are 200 mm x 200 mm.*

Effects of using a fractal-based ground plane

The fractal-based ground plane is created using four copies of a Sierpinski gasket rotated 90º to each other. In the ground plane generation process a different number of fractal iterations can be used obtaining the geometries shown in figure 2.15. In this figure it can be seen that the generated ground plane keeps the self similarity property of the original Sierpinski gasket, and hence it is predictable that the ground plane will also exhibit multiband behavior with multiple controlled resonances.

Figure 2.16 shows that new resonances are introduced when the number of ground plane fractal iterations is increased. This is because the whole structure resonances are a combination of the monopole element resonances and the ground plane resonances. Moreover, the position of the ground plane resonances is similar to the position of the resonances obtained if the Sierpinski gasket used in the ground plane generation process is used as monopole element. Only a small frequency shift is observed due to the mutual loading effects between the different copies of the Sierpinski gaskets that make up the ground plane. Furthermore, good matching levels are obtained in the odd resonance when

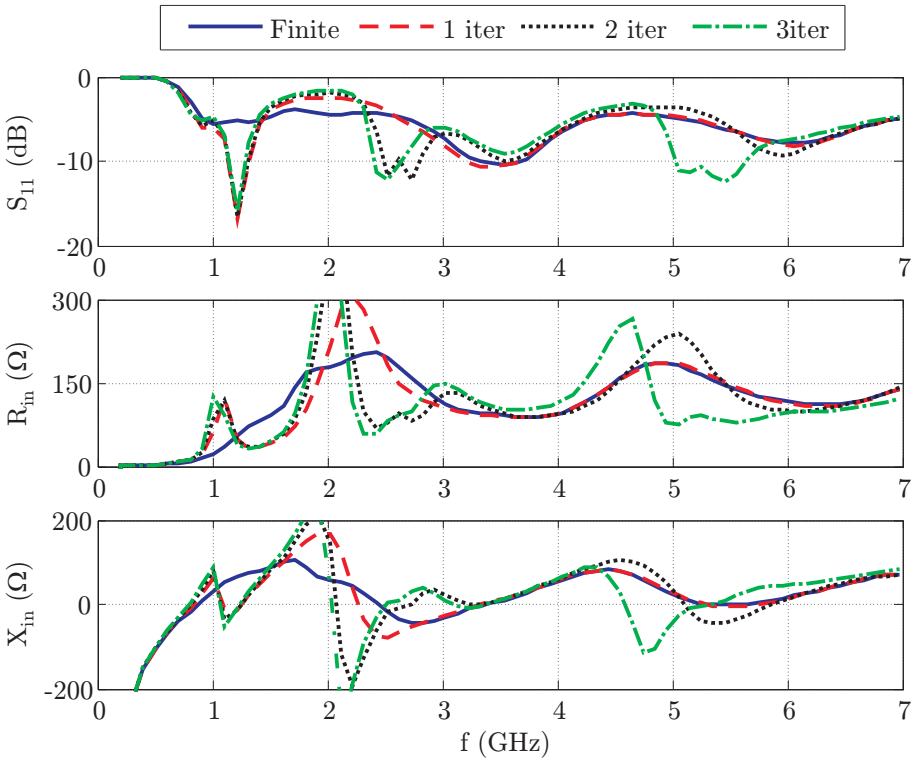


Figure 2.16: *Input parameters of a triangular monopole on a fractal-based ground plane with different number of iterations (figure 2.15).*

the reference impedance is 50 Ω. It is important to point out that the aforementioned perturbations applied to the Sierpinski monopole, like the use of different scale factors, can also be applied to the ground plane in order to adjust the position of its resonances.

Dual band triangular monopole antenna on a fractal-based ground plane

In this section an equilateral triangular monopole will be mounted on a fractal-based ground plane (figure 2.17) to obtain a multiband antenna that covers the needs of the central unit.

The design process is as follows. The triangular monopole height h_m is set in order to obtain 50 Ω matching in the lower part of the 5 GHz band when it is on a solid ground plane. In the generation of the Sierpinski gasket for the ground plane, two fractal iterations with a scale factor of 2 were initially used, with the objective of obtaining three operating bands with frequency spacing around 2. However, the spacing between the two lower bands is larger than 2 due to the truncation effects of the real finite ground plane. Moreover, the first band is not well matched to 50 Ω, therefore, this band will not be considered in our design, setting $h_1 = 2h_2$ just to avoid the truncation effect (mismatch) for the second

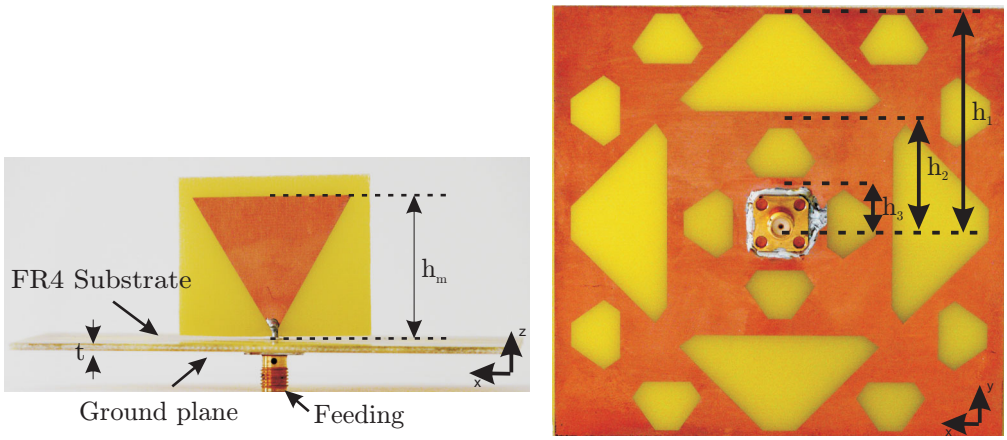


Figure 2.17: *Triangular monopole on a fractal-based ground plane. (a) Monopole element. (b) Modified ground plane inspired by the Sierpinski gasket.*

band.

The heights h_2 and h_3 have to be adjusted to allocate the 2.45 GHz and 5 GHz WLAN bands. For this reason the scale factor of the second iteration is increased to 2.36 to obtain the desired frequency spacing. The final design (figure 2.17) has the following dimensions: $h_m = 30$ mm, $h_1 = 52$ mm, $h_2 = 26$ mm, and $h_3 = 11$ mm. These dimensions take into account the effects of the low cost FR4 1.55 mm substrate ($\epsilon_r = 4.5$, $tan\delta = 0.02$) used to support the antenna metallic parts.

In order to demonstrate the improved performance of the antenna with the modified ground plane, it will be compared with the results obtained when the triangular monopole is mounted on a solid ground plane with the same dimensions. The input impedance and return losses of both configurations are plotted in figure 2.18, and table 2.1 summarizes the operating bands and their bandwidth, taking into account that an input reflection coefficient better than -10 dB is desired.

For the triangular monopole on the solid ground plane there are two possible operating bands. However, the matching levels of the first one are poor and therefore it may not be useful. The second one, with a center frequency of 5.1 GHz, is well matched and has a relative bandwidth of 17.2 %. In the case of the fractal-based ground plane, new resonances appear due to the fractal shape, obtaining good 50 Ω matching at 2.475 GHz, with a relative bandwidth of 6.3 %. Moreover, the 5 GHz band has been improved, increasing its relative bandwidth to 28.6 %. This is due to the fact that dimensions have been adjusted to obtain the third resonance introduced by the ground plane at a frequency slightly higher than the monopole resonance. The combination of both resonances is responsible for the

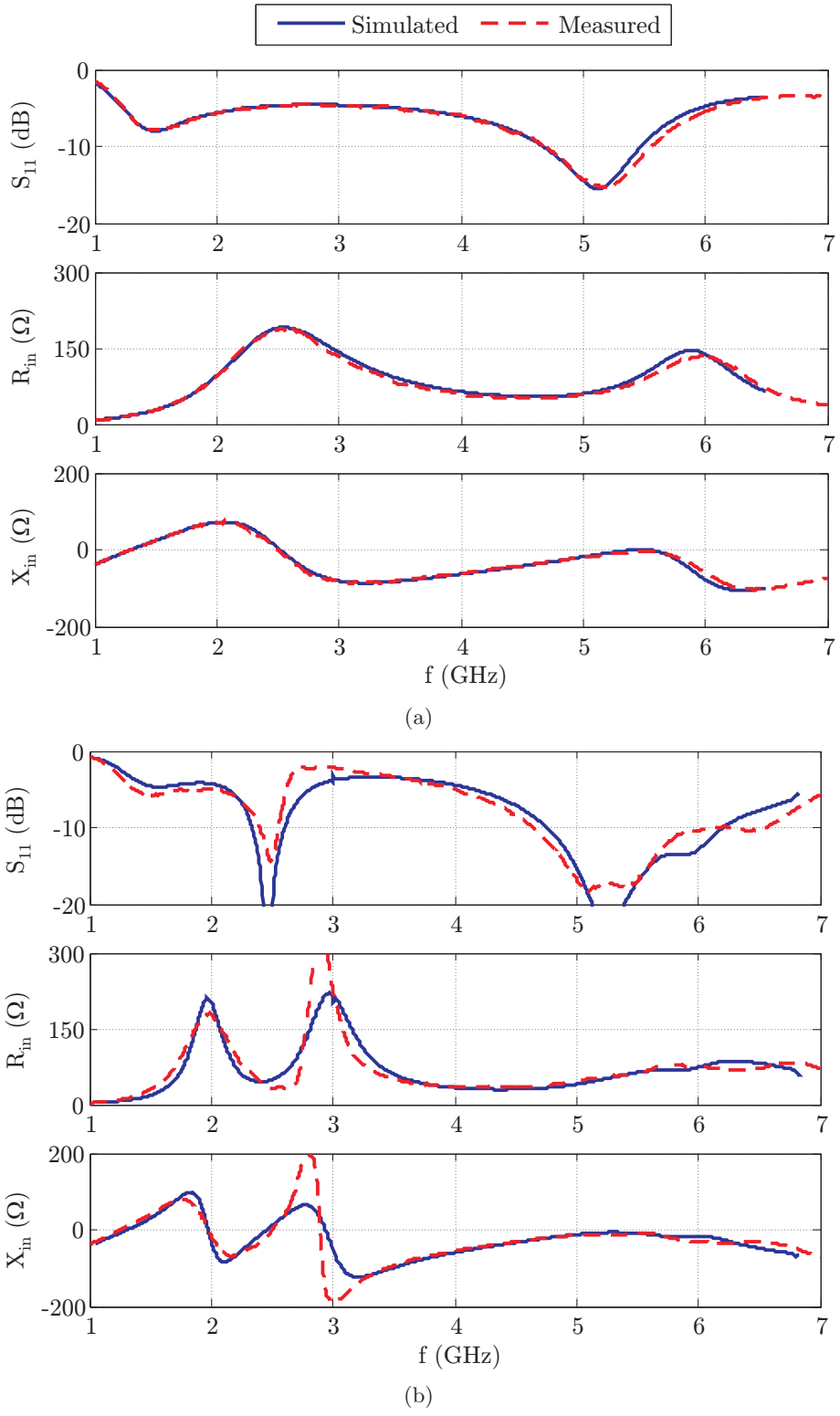


Figure 2.18: *Antenna input parameters. (a) Triangular monopole on a solid ground plane. (b) Triangular monopole on a fractal-based ground plane.*

Table 2.1: *Measured operating bands of the triangular monopole on the different ground plane configurations.*

Solid ground plane		
Operating band (GHz)	Center frequency (GHz)	Relative bandwidth(%)
4.675 - 5.556	5.1	17.2

Fractal-based ground plane		
Operating band (GHz)	Center frequency (GHz)	Relative bandwidth(%)
2.397 - 2.552	2.475	6.3
4.645 - 6.194	5.4	28.6

bandwidth increase.

The simulated copolar and crosspolar components of both configurations are shown in figure 2.19. Good similarity between the two bands and reduced crosspolar levels are obtained. The radiation patterns keep the same properties when the solid ground plane is replaced by the fractal-based ground plane, proving that the fractal ground plane allows the introduction of new operating bands without significantly degrading the radiation parameters of the original monopole antenna. The simulated directivity with the solid ground plane is 3.4 dB for the first band and 4.6 dB for the second one, whereas the directivity with the fractal-based ground plane is 4.1 dB for the first one and 5.1 dB for the latter.

Back radiation could be considered significant. However, this is not due to the use of a fractal-based ground plane but to the small ground plane dimensions, because similar back radiation levels are obtained with both ground plane configurations. These back radiation levels could be improved by using larger ground planes. The simulated radiation efficiency is over 90% in the frequency range under consideration for both cases.

Figure 2.20 shows the current distributions in the fractal-based ground plane in both operating bands. It can be seen as frequency increases, currents are located in a smaller region of the ground plane. Due to the self similarity property of the fractal geometry, current distributions at the different bands are similar with different scale factors, and this allows us to obtain the multiband performance of the antenna.

This antenna offers good matching levels, good similarity in radiation patterns and good efficiency levels in both frequency bands, and thus it could be perfectly well used as antenna for the central unit.

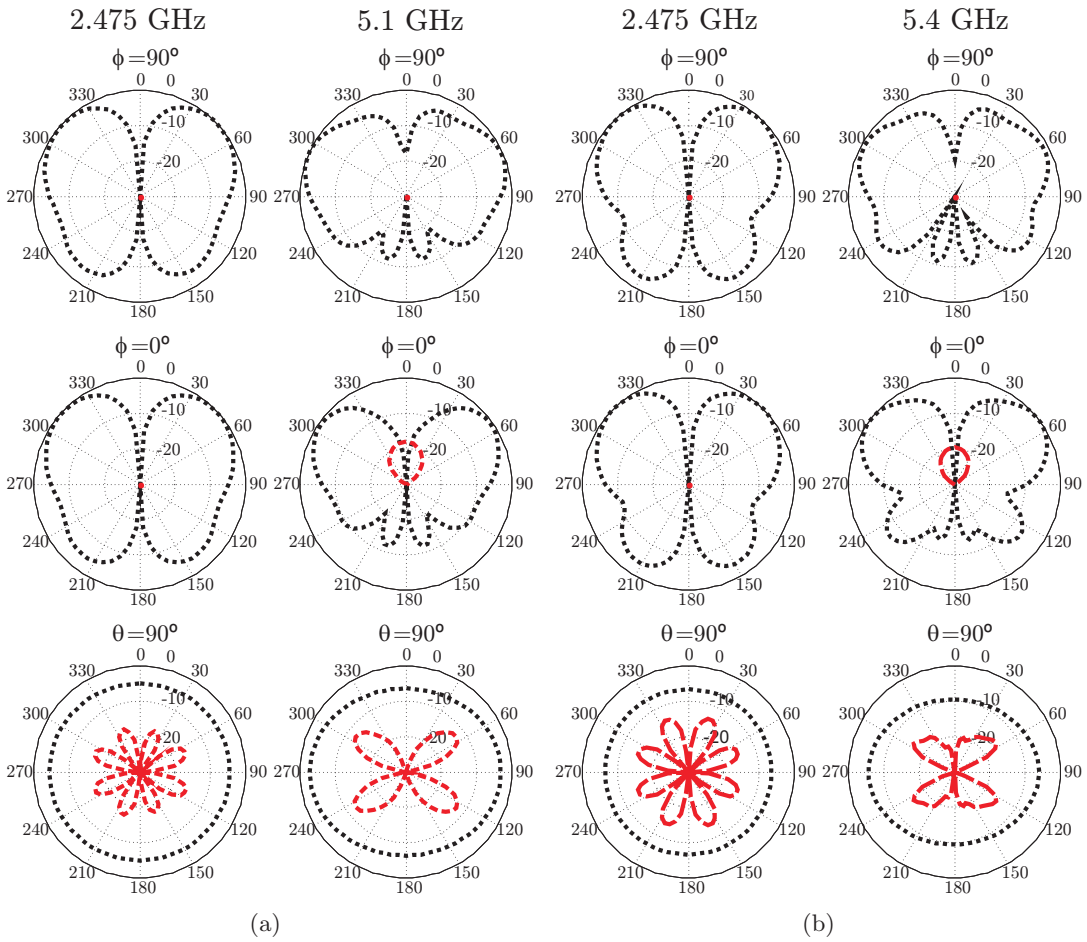


Figure 2.19: *Simulated normalized radiation patterns. Black dotted lines are used for the copolar component, red dashed lines are used for the crosspolar component. (a) Triangular monopole on a solid ground plane. (b) Triangular monopole on a fractal-based ground plane.*

Dual band Sierpinski monopole antenna on a fractal-based ground plane

In the previous section the fractal shape was only used in the ground plane in order to introduce new operating bands to a triangular monopole. In this section, the Sierpinski gasket will be used in both elements to increase design flexibility. The fractal-based ground plane will reinforce the resonances of a Sierpinski monopole improving the matching levels and bandwidths.

The ground plane is similar to the one presented in figure 2.17(b), but dimensions are slightly different with $h_1 = 50$ mm and a scale factor equal to two. The monopole is based on a Sierpinski gasket with the same scale factor. As can be seen in figure 2.21, several monopoles with various flare angles α were fabricated.

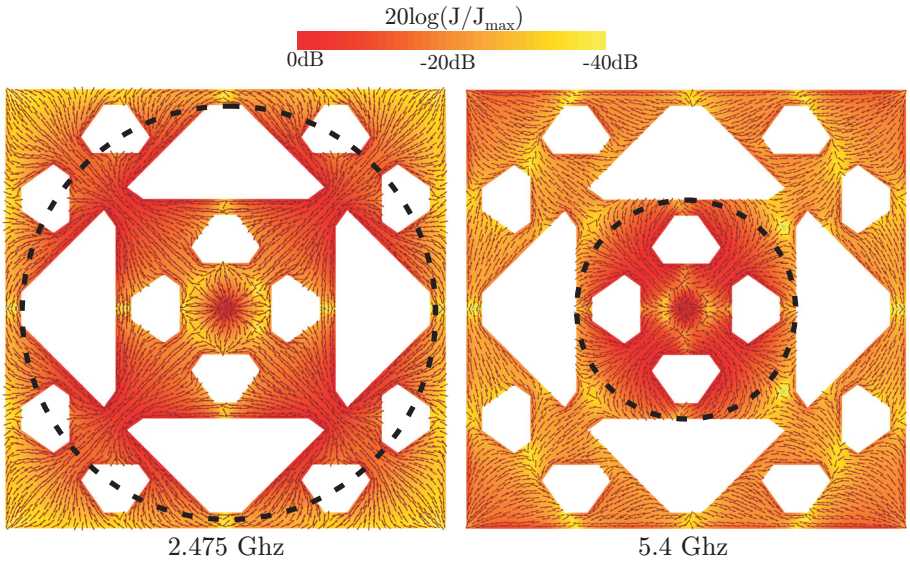


Figure 2.20: *Current distributions in the ground plane. The area with higher current density levels (active area) has been marked with a dashed circle.*

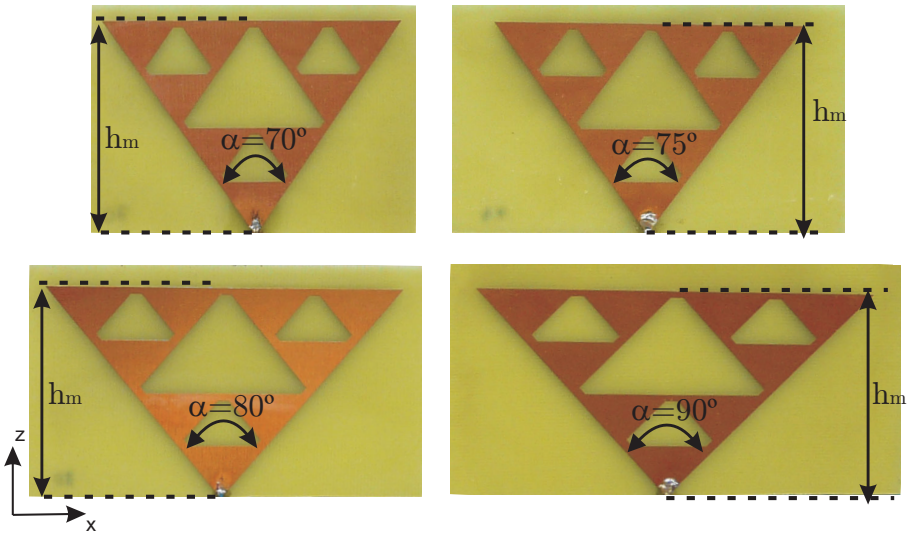


Figure 2.21: *The monopole element is a two iteration Sierpinski gasket with different flare angles α. Antennas were fabricated on a low cost FR4 substrate (ϵ_r = 4.5, tanδ = 0.02). h_m = 50 mm.*

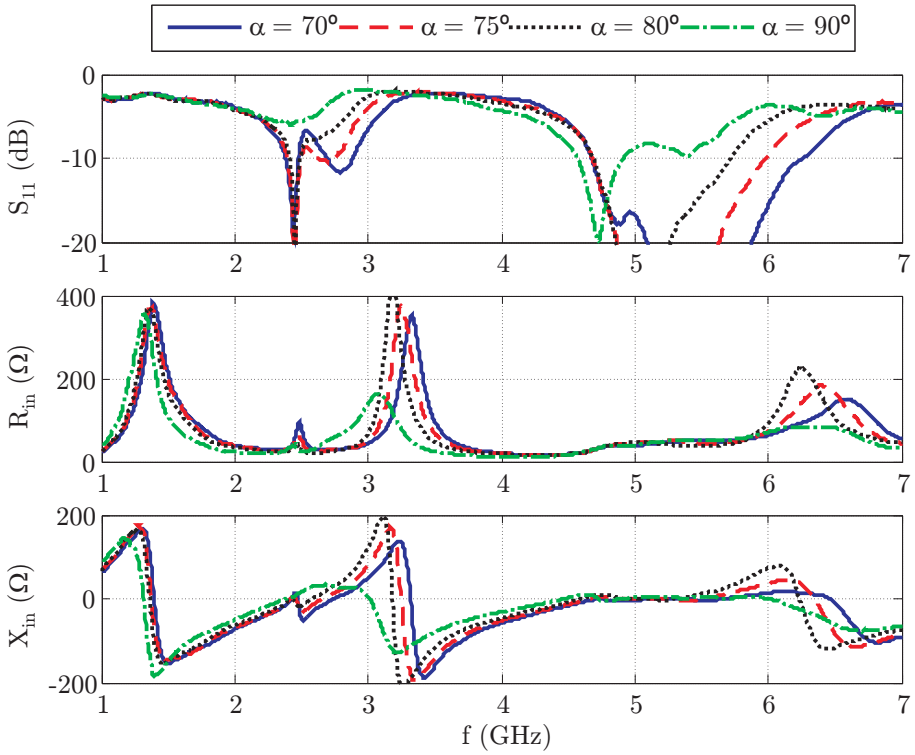


Figure 2.22: *Measured input parameters of a Sierpinski monopole with different flare angles α on a fractal-based ground plane.*

If the Sierpinski gasket used as monopole and the Sierpinski gasket used in the ground plane generation process are the same, both will have approximately the same resonance frequencies, reinforcing the resonances of the whole structure. However, the antenna performance can be improved if the monopole is modified in order to separate the resonances of both elements and increase the available bandwidth.

Antenna measurements of figure 2.22 show that double resonances appear near the operating bands when the monopole flare angle is not 90º. Since the flare angle α of the monopole is different from the flare angle of the Sierpinski gasket used in the ground plane, resonance frequencies of both elements are not the same, creating double resonances in the response of the whole antenna that help to improve the bandwidths. Similar results could be obtained by changing the monopole height h_m instead of the flare angle α.

As the flare angle decreases, monopole resonance frequencies slightly increase and the frequency separation between resonances also increases improving the bandwidth. The problem is that if the monopole flare angle is smaller than $\alpha = 75^{\circ}$, the monopole resonances and the fractal-based ground plane resonances are separated too far and the first band (2.45 GHz) splits into two bands with a reduced bandwidth.

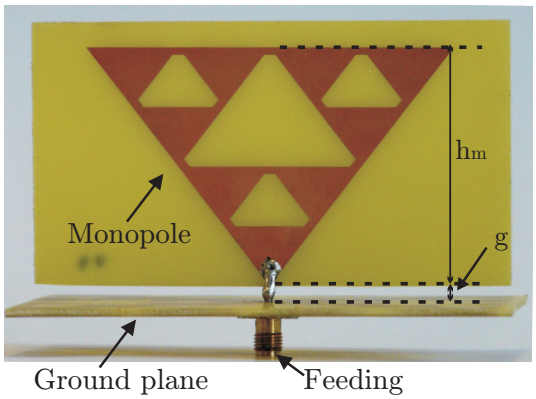


Figure 2.23: *A small gap g between the ground plane and the monopole can be used to optimize the antenna performance and improve the matching levels.*

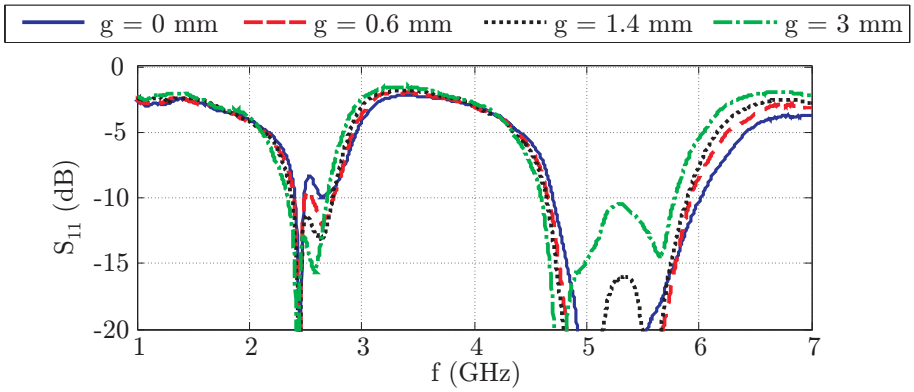


Figure 2.24: *Measured input parameters of a Sierpinski monopole with $\alpha = 75^o$ mounted on a fractal-based ground plane with different gap sizes g (figure 2.23).*

Once the flare angle is set to $\alpha = 75^o$, a fine tuning process to optimize the matching levels can be carried out by adjusting the height of the monopole (g) above the ground plane (figure 2.23). Measured return losses for different values of g are plotted in figure 2.24. As the height g increases, the insertion losses at the first band decrease while the second band is slightly degraded. The optimum choice of g in order to maximize the $S_{11} < -10$ dB bandwidth in both bands is $g = 1.4$ mm. With this configuration the operating bands are 2.37 GHz - 2.74 GHz and 4.64 GHz - 5.9 GHz, with relative bandwidths of 14.5 % and 23.9 % respectively. Table 2.2 summarizes the information about the operating bands with and without the fractal-based ground plane.

If we compare these results with those presented in the previous section where a triangular monopole was used, we can see that the use of the fractal shape simultaneously in the monopole and in the ground plane allows us to increase the first band bandwidth without significantly degrading the second band.

Table 2.2: *Measured operating bands of the Sierpinski monopole on the different ground plane configurations.*

Solid ground plane		
Operating band (GHz)	Center frequency (GHz)	Relative bandwidth(%)
2.58 - 2.73	2.65	5.6
4.75 - 5.23	5	9.6

Fractal-based ground plane		
Operating band (GHz)	Center frequency (GHz)	Relative bandwidth(%)
2.37 - 2.74	2.55	14.5
4.64 - 5.90	5.27	23.9

Simulated copolar and crosspolar components of the radiation patterns for the optimized design are shown in figure 2.25(b). For comparative purposes, figure 2.25(a) shows the radiation patterns obtained when the same Sierpinski monopole is mounted on a solid ground plane with the same dimensions as the fractal ground plane. By comparing figures 2.25(a) and 2.25(b) it can be concluded that the fractal ground plane does not introduce important changes in the radiation patterns and that low crosspolar levels are obtained. The simulated directivity is around 4.2 dB for the first band and 5.8 dB for the second. Although low cost FR4 dielectric losses are high, simulated efficiency levels are over 70 % in both bands.

As in the previous section, the presented antenna offers good multiband performance, and it could be used for the central unit.

2.3 Resonator loaded multiband monopole antennas

In this section, a new kind of antenna suitable for the central unit will be analyzed: monopoles loaded with resonators. The resonators effects will be used to improve the antenna multiband behavior.

2.3.1 Introduction to resonator loaded antennas

Recently, it has been demonstrated that loading a dipole/monopole with 'particles' can also facilitate dual band operation [64, 65]. These 'particles' are resonators that couple via the magnetic field produced by currents flowing along the dipole. In effect, such resonators load the dipole element at specifically chosen locations with a reactance that is determined by the resonant properties of the particular 'particle' geometry. The effects of these resonators

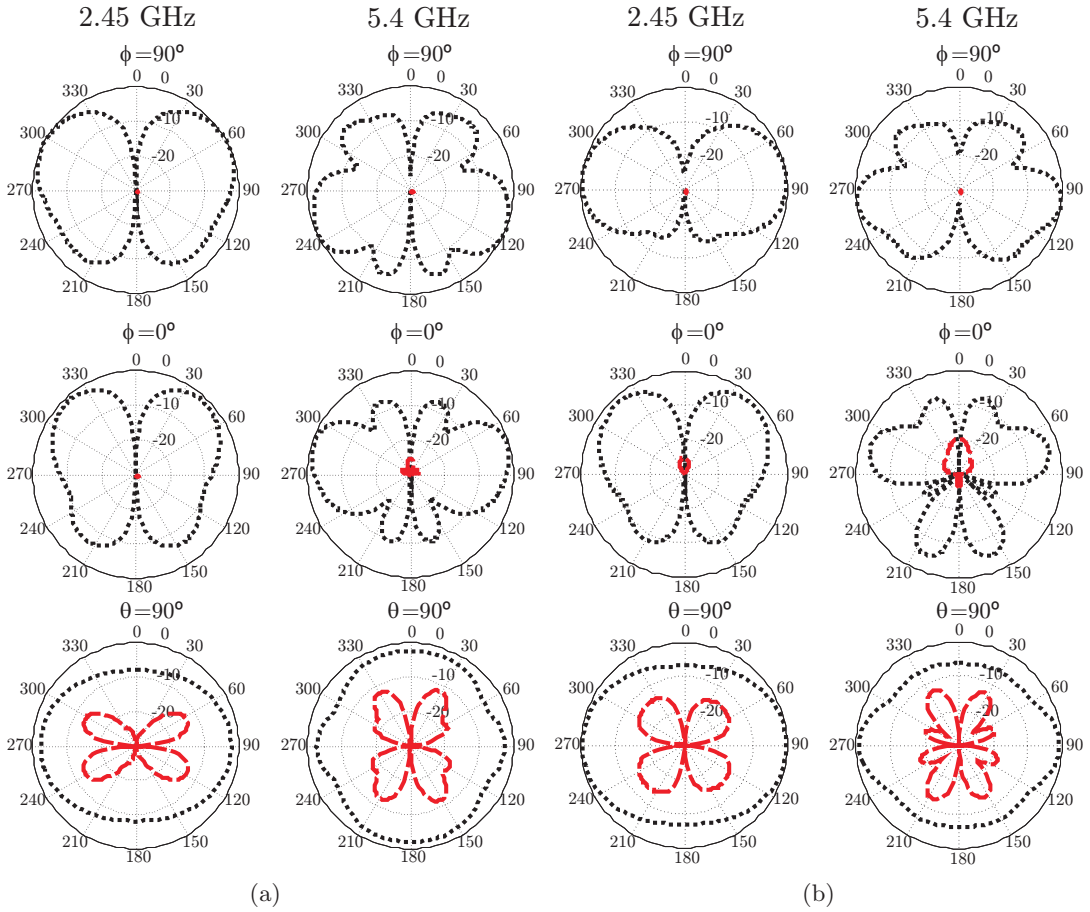


Figure 2.25: *Simulated normalized radiation patterns. Black dotted lines are used for the copolar component, red dashed lines are used for the crosspolar component. (a) Sierpinski monopole on a solid ground plane. (b) Sierpinski monopole on a fractal-based ground plane.*

can introduce new matching bands to the antenna.

Amateur radio enthusiasts have used a similar principle in trap antennas operating in HF bands for many years [66]. For operating in a number of bands, such as those between 3.5 MHz and 30 MHz, it is impractical for most amateurs to use a separate antenna for each band, for this reason multiband antennas are also necessary in this frequency range. As an example, in [66] a trap antenna with the structure shown in figure 2.26 for the 3.5 MHz and 7 MHz bands is presented.

A trap in an antenna can perform two different functions depending on whether or not it is resonant at the operating frequency. A typical use is a resonant parallel trap. Let us assume that each LC combination is resonant in the 7 MHz band. Because of this parallel resonance, the trap presents high impedance at this point of the antenna. The electrical

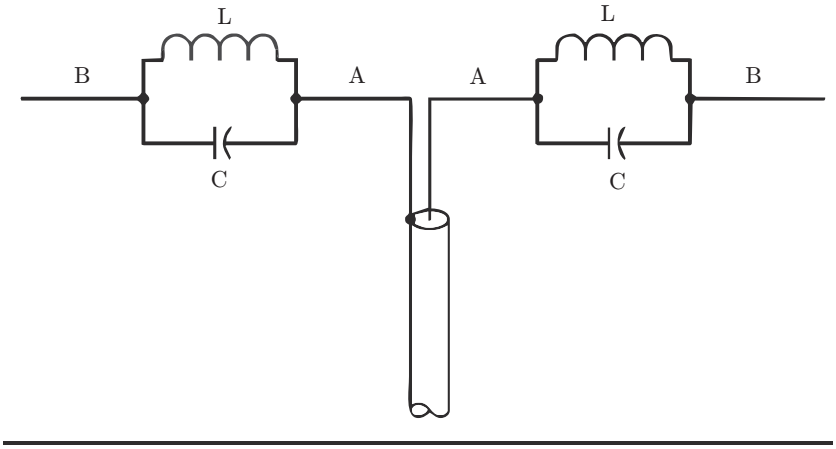


Figure 2.26: *Schematic representation of a trap dipole antenna [66].*

effect at 7 MHz is that the trap behaves as an insulator, an open circuit that 'cuts' the antenna at this point, separating the B sections from the antenna. The result is that we have an antenna system that is resonant in the 7 MHz band, taking into account that each section A represents $\lambda/4$ at this frequency. One of the problems of this configuration is that the trap losses at the resonance frequency can reduce the antenna efficiency.

At the other operating band, where the trap is not resonant, the LC resonators acts as an electric load. If the operating frequency is below the trap resonance, the trap acts as an inductor; if above, as a capacitor. Inductive loading will electrically lengthen the antenna, and capacitive loading will electrically shorten the antenna. In the example we are considering, the traps behave as inductors when operation takes place at 3.5 MHz, electrically lengthening the antenna. This means that the total length of sections A and B may be something less than a physical $\lambda/4$ at 3.5 MHz. Thus, we have a dual band antenna that is shorter than a resonant dipole in the lower frequency band.

In the previous example, traps were resonant at one of the two frequencies of operation. However, the same band coverage can be obtained with resonant traps at a frequency between both operating bands, for instance, 5 MHz. With proper selection of the LC ratio and the dimensions of A and B sections, the traps will act to electrically shorten the antenna at 7 MHz and electrically lengthen it at 3.5 MHz. Thus, an antenna that is intermediate in physical length between a resonant dipole at 3.5 MHz and a resonant dipole at 7 MHz can cover both bands, even though the trap is not resonant at these frequencies. Note that these traps have less RF current flowing in the trap components in the operating bands, and hence trap losses are lower. The use of a trap resonating at a frequency quite apart from the antenna operating bands is also known as off-resonance loading.

Based on these concepts and operating principles, different antenna topologies with multiband behavior suitable for the central unit of the sensor network will be presented and we will also give some guidelines for the antenna design process. In the frequency bands used by the sensor network, traps will be implemented in a distributed way instead of using discrete components, introducing some advantages such as the fact that the additional physical path created by the resonators can be used to tune the position of the straight monopole resonances.

2.3.2 Dual band resonator loaded monopole antenna

In this section the off-resonance loading principle will be employed in a design that avoids resonator losses using magnetic resonators whose natural frequency lies between the lower and upper WLAN bands.

Antenna design

The basic geometry of the antenna consists of two identical capacitively loaded magnetic resonators attached to a uniform width monopole (figure 2.27). The magnetic resonators

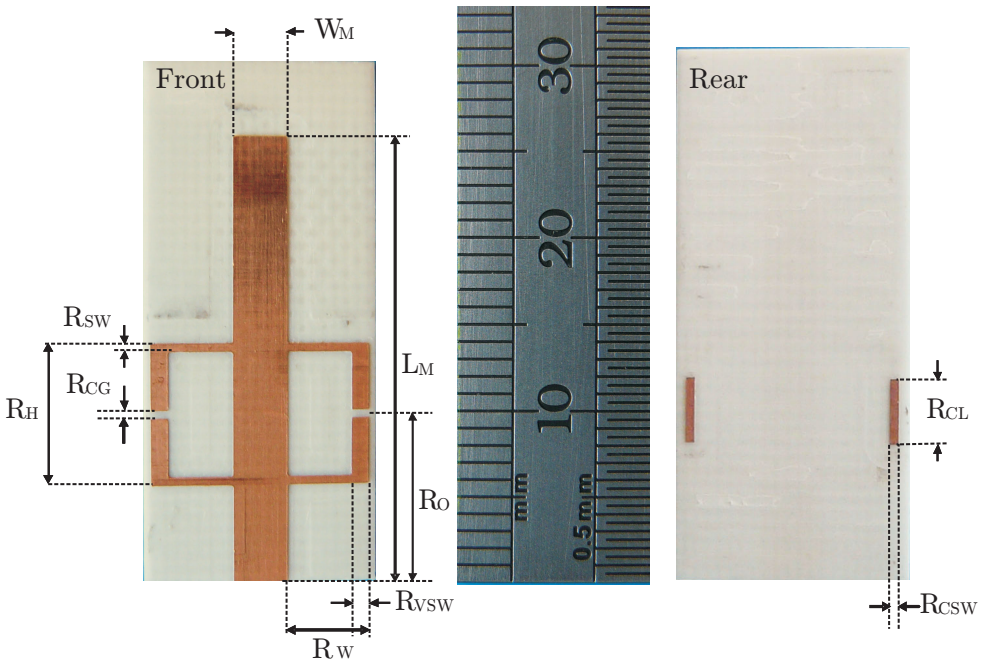


Figure 2.27: *Monopole antenna loaded with two double layer magnetic resonators DL-CLL [67] of identical dimensions fabricated on Rogers RO4003C [68] double sided 0.508 mm substrate (ϵ_r = 3.38, tanδ = 0.0021).*

are based on a double layer capacitive loaded loop (DL-CLL) [67].

The monopole length L_M affects both operating bands (figure 2.28(a)), so the design begins with L_M approximately equal to a quarter wavelength at 2.45 GHz. A small adjustment of this parameter must then be carried out to fine tune the first band position.

According to figure 2.28(b) the resonator width R_W simultaneously shifts the second and the third S_{11} minimums without significantly affecting the first. The second S_{11} minimum shifts because the resonator inductance changes with R_W. The third S_{11} minimum position is also affected by the resonator width because, as will be shown in figure 2.31, at frequencies near this S_{11} minimum currents tend to follow the perimeter

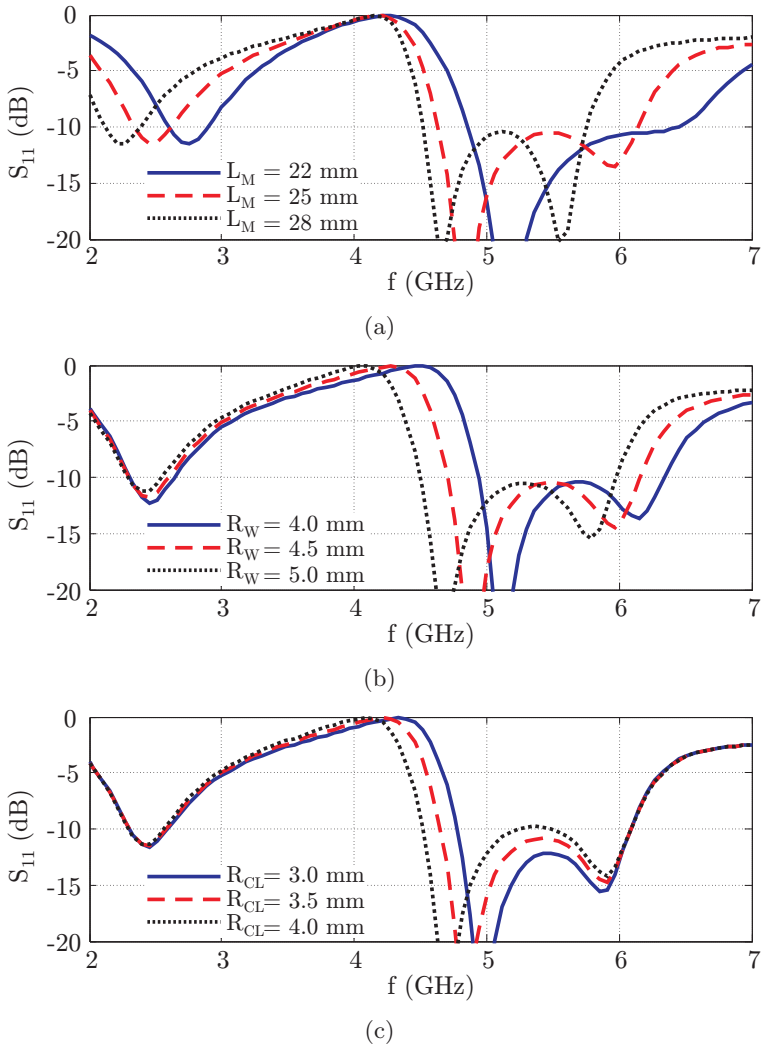


Figure 2.28: *Parametric study of the variation of the input reflection coefficient S_{11} with respect to (a) the monopole length L_M, (b) the resonator width R_W, (c) the length of the resonator capacitor strip R_{CL}.*

of the structure. Consequently, the resonator external width R_W is chosen based on the wavelength λ_2 corresponding to the upper frequency of the second band defined by the 802.11 standard (5.8 GHz) such that

$$L_M + 2R_W \approx \frac{3\lambda_2}{4} \tag{2.2}$$

The second S_{11} minimum is mainly controlled by the resonance frequency of the resonator (figure 2.28(c)). The resonator dimensions are set to obtain a resonance frequency lower than 4.9 GHz. This can be achieved by changing the resonator height R_H (modifying the inductance), the loop gap R_{CG}, and the rear strip dimensions R_{CL}, R_{CSW} (modifying the capacitance). The separation between the second and the third S_{11} minimums affects the second band bandwidth and matching levels. A clear trade-off between both parameters exists: when the separation is increased, bandwidth is improved, but center band matching is degraded.

The resonator quality factor will also affect the bandwidth of the impedance match over the 5 GHz band. The bandwidth can be increased by decreasing its capacitance and increasing the resonator height R_H in order to increase the inductance. The resonator loop strip width R_{SW} is chosen as 0.4 mm based on fabrication considerations.

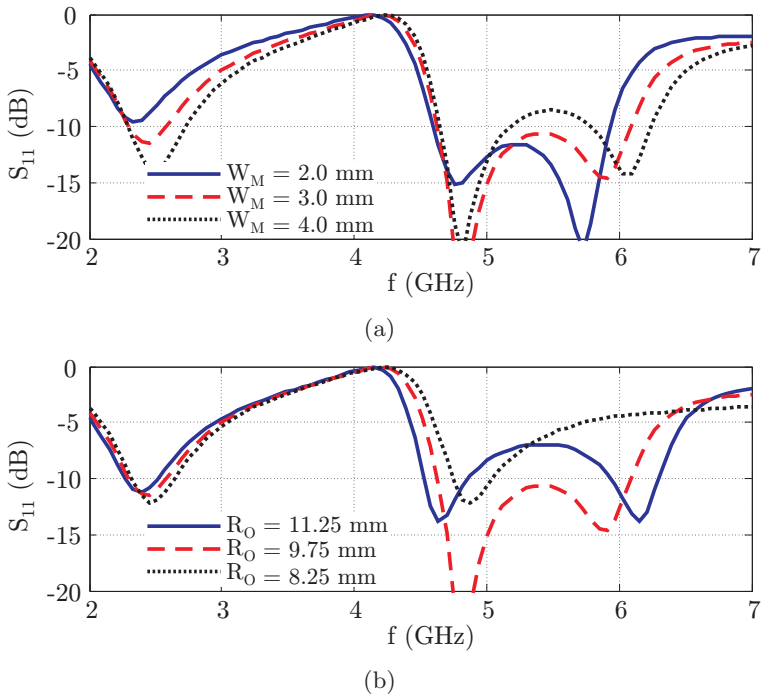


Figure 2.29: *Parametric study of the variation of the input reflection coefficient S_{11} with respect to (a) the monopole width W_M and (b) the resonator offset position R_O.*

The match over both operating bands can be optimized by adjusting the monopole width W_M and the resonator position offset R_O (figure 2.29(a) and 2.29(b)). Furthermore, the offset parameter controls the resonator loading position, and the adjustment of this parameter allows for the separation of the minimum efficiency point, due to resonator losses at the resonance frequency, from the match band (off-resonance loading principle).

After the initial design, a fine tuning process of the aforementioned parameters is necessary to allocate the operating bands according to the system requirements.

Prototype antenna: Fabrication and measurement

Figure 2.27 shows the prototype antenna fabricated on Rogers RO4003C [68] 0.508 mm substrate ($\epsilon_r = 3.38$, $tan\delta = 0.0021$) using a LPKF ProtoMat S62 circuit board plotter. Prototype dimensions are shown in table 2.3. The capacitor formed by overlapping strips on both faces is the most tolerance critical part of the fabrication process. The alignment tolerance was relaxed by making the front strip twice as wide as the rear strip. The prototype presented here is the second of two that were built on double-sided substrate, the first one covered 802.11 a/b/g/h WLAN bands and the second one, the universal 802.11 a/b/g/h/j WLAN bands that include Japan.

The input reflection coefficient of the fabricated prototype is plotted in figure 2.30. The available relative bandwidths (table 2.4) are enough to cover the WLAN and ISM bands requirements. In the 2.45 GHz band the relative bandwidth is 11.7 %. In the 5 GHz band, the combination of the second resonance of the straight monopole near 6 GHz and

Table 2.3: *Dimensions of the dual band WLAN monopole antenna.*

L_M	W_M	R_{SW}	R_W	R_H
25.5 mm	3 mm	0.4 mm	4.68 mm	8 mm

R_{VSW}	R_{CG}	R_O	R_{CL}	R_{CSW}
0.9 mm	0.5 mm	9.75 mm	3.6 mm	0.45 mm

Table 2.4: *Measured operating bands of the dual band resonator loaded monopole antenna (figure 2.27) with the dimensions of table 2.3.*

Resonator loaded monopole antenna		
Operating band (GHz)	Center frequency (GHz)	Relative bandwidth(%)
2.26 - 2.54	2.4	11.7
4.74 - 6.12	5.4	25.4

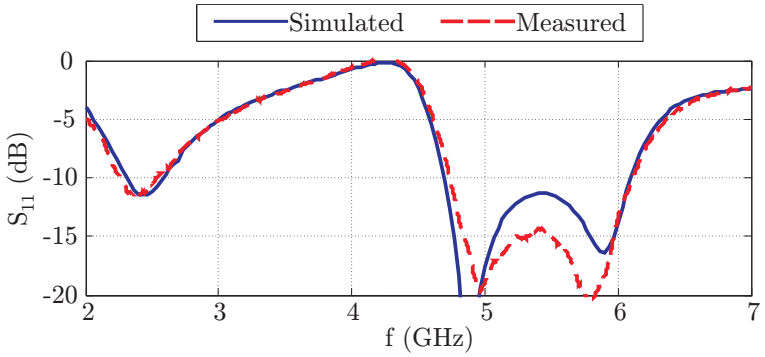


Figure 2.30: *Simulated and measured S_{11} parameters of the prototype antenna with the dimensions shown in table 2.3.*

the S_{11} minimum introduced by the resonator effects near 5 GHz, is used to improve the available bandwidth, achieving a relative bandwidth of over 25 %.

The normalized current distributions (figure 2.31) show a $\lambda/4$ standing wave at 2.45 GHz where currents in the resonator arms are practically zero, a $\lambda/4$ standing wave at 4.9 GHz where the resonator effectively isolates the top half of the monopole, and a $3\lambda/4$ standing wave at 5.875 GHz over a path that includes the monopole top and bottom strips and the resonator arms. This explains why the first S_{11} minimum position is mainly defined by the monopole length, the second S_{11} is mainly defined by the properties of the

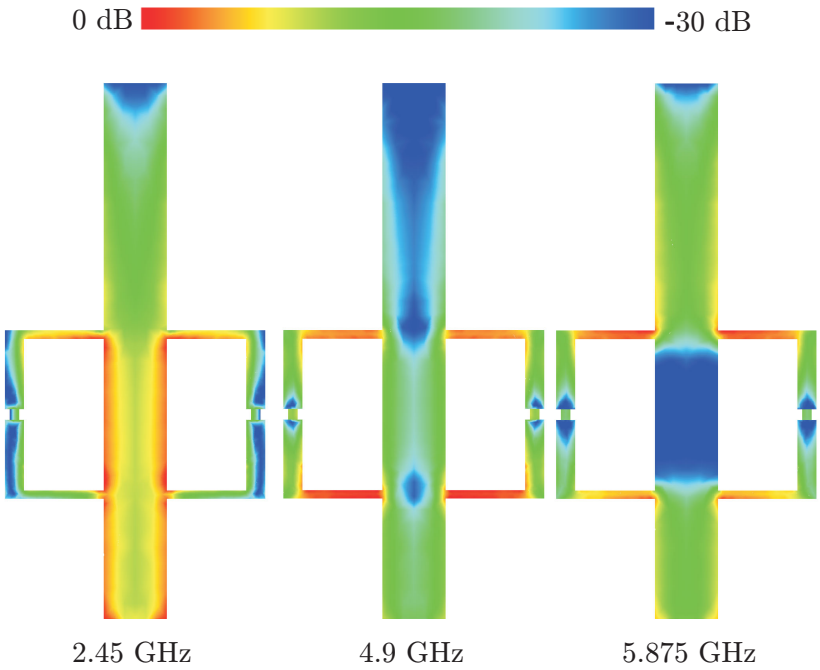


Figure 2.31: *Simulated current distributions at 2.45 GHz, 4.9 GHz and 5.875 GHz. Each image is individually normalized and displayed on a 30 dB scale.*

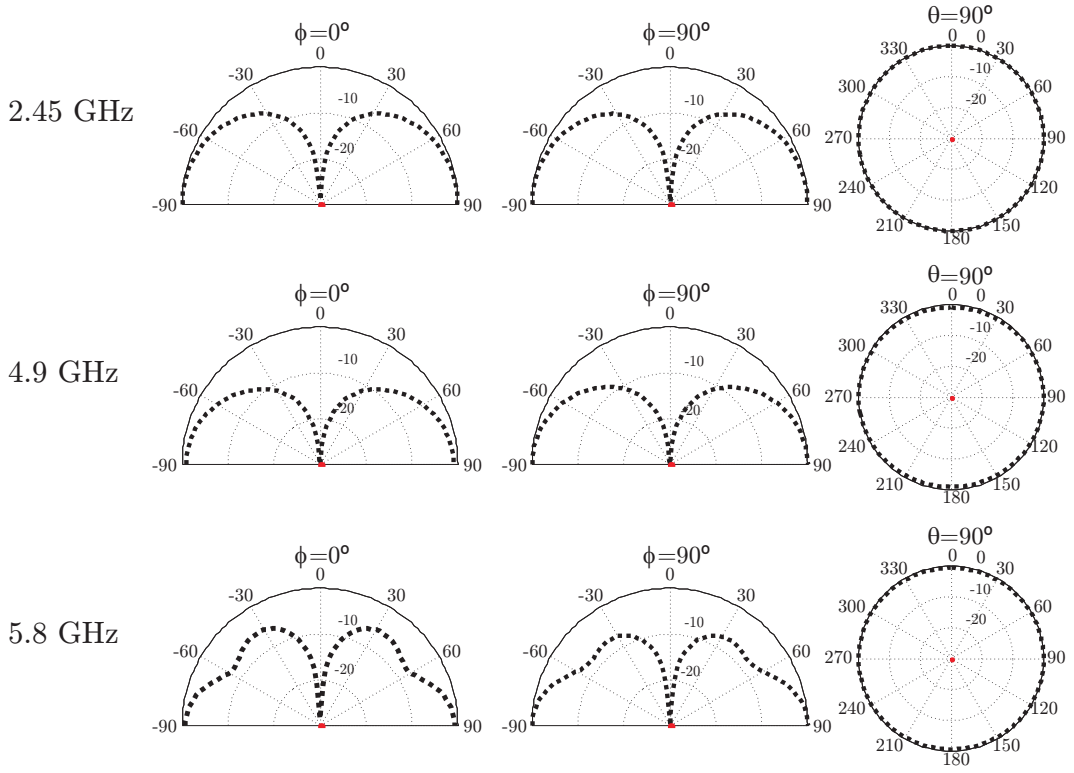


Figure 2.32: *Simulated radiation patterns of the prototype antenna whose dimensions are shown in table 2.3. Black dotted lines are used for the copolar components. Crosspolar components are too low to be plotted with the current axis configuration.*

monopole and the resonator, and the third S_{11} minimum is mainly defined by the length of the external perimeter of the structure (including the resonator arms).

Figure 2.32 shows the radiation patterns.There is little variation in the shape of the 2.45 GHz radiation patterns, so only one is shown in this frequency band. The patterns at 4.9 GHz are characteristic of a quarter wavelength monopole, while, as expected, those at 5.8 GHz show a three quarter wavelength monopole pattern. The H-plane radiation patterns are essentially omnidirectional with less than 1 dB variation over both bands. Due to the antenna symmetry, the simulated maximum crosspolar levels are exceptionally low. The gain ranges from a maximum of 7.1 dB to a minimum of 5.8 dB.

The radiation and total efficiencies of the fabricated prototype are plotted in figure 2.33. Measurements were carried out using the method presented in [69] and FEKO [70] was used for the simulations. The minimum efficiency occurs at the resonator resonance of 4.25 GHz, some 650 MHz below the band edge frequency of 4.9 GHz. The in-band radiation efficiency is practically unaffected by the resonators thanks to the use of the off-resonance loading principle where the resonator resonance is out of the antenna operating bands. Measured

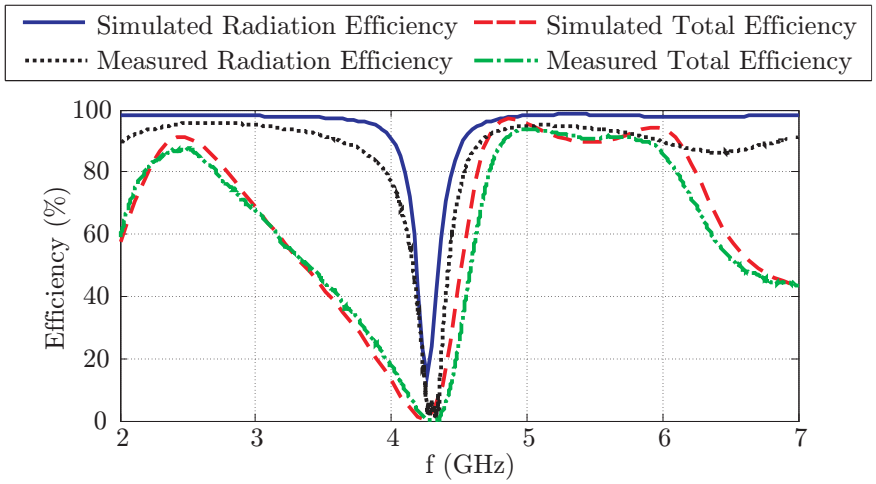


Figure 2.33: *Radiation efficiency and total efficiency of the prototype whose dimensions are shown in table 2.3. Efficiency measurements were carried out using the method presented in [69].*

total efficiency is higher than 85 % in both operating bands. For technical reasons, Method of Moments - Finite Element Method (MoM-FEM) FEKO simulations include dielectric losses but exclude copper losses, resulting in optimistic in-band radiation efficiency and somewhat larger errors out-of-band near the resonator resonance where copper losses are particularly important. Interestingly, the measured efficiencies confirm the overall FEKO predictions, which are especially useful for determining that the frequency of the minimum radiation efficiency is out-of-band.

The novel antenna analyzed in this section is also suitable for the central unit. Furthermore it is important to point out that its dimensions could be reduced loading the monopole with only one resonator, however with this asymmetric configuration the antenna performance is slightly degraded and crosspolar levels are increased.

2.3.3 Triple band SRR loaded monopole antenna

In the previous section a resonator loaded monopole has been presented in order to obtain a dual band radiator. Now, we will present a resonator loaded monopole that takes advantage of the dual band behavior of the Split Ring Resonator (SRR) to achieve triple band performance.

The SRR has already been used to load a dipole and obtain dual band behavior [64, 65]. In [65] the lower band was defined by the dipole resonance whereas the upper band was based on the first SRR resonance. However, the SRR also exhibits an upper resonance which has led recent papers to propose these 'particles' as basic building blocks for dual

band filters [71, 72]. Therefore we will present a structure that combines the dual band SRR with the first resonance of a monopole to obtain triple band performance with good radiation efficiency and similar radiation patterns in all the three bands. In this design, the new frequency band will be used to cover one band of the 802.15 standard (3.244 - 4.742 GHz).

SRR analysis

Conventional SRRs are made up of two concentric open rings with their slits on opposite sides. To evaluate the resonance frequency of the isolated rings and the SRR in free space we followed the procedure presented in [73] and calculated the monostatic radar cross section of the structure when a plane wave impinges as shown in figure 2.34.

Because of the coupling between rings, the SRR has two resonances (figure 2.35): one below and one above the individual resonances of the isolated rings. The SRR resonances depend on the length of the open loops and the gap between rings which controls the strength of the coupling. More details about the SRR design can be found in [71].

The simulated SRR current distributions (figure 2.36) agree with those in [71]: in the lower resonance the induced circulating current generates a magnetic dipolar moment aligned with the impinging magnetic field, whereas an electric dipolar moment aligned with the impinging electric field is created in the upper resonance.

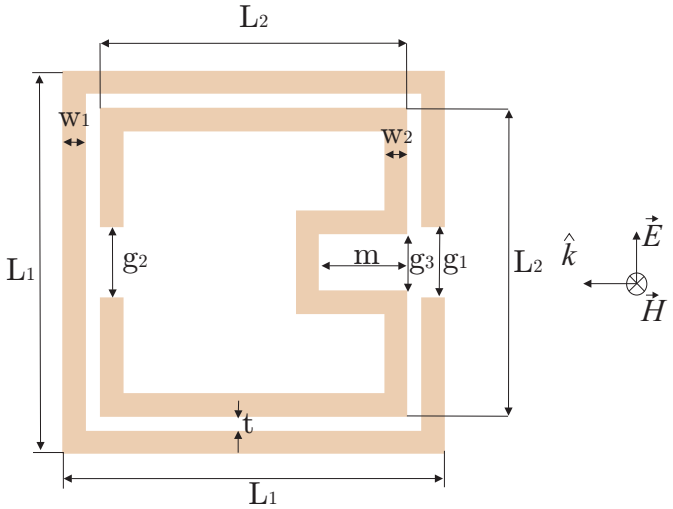


Figure 2.34: *SRR geometry and impinging plane wave for computation of the monostatic radar cross section. The SRR is supported on Rogers RO4003C 0.508 mm substrate ($\varepsilon_r = 3.38$, $\tan\delta = 0.0027$). $L_1 = 8.2$ mm, $L_2 = 6.6$ mm, $w_1 = w_2 = 0.5$ mm, $g_1 = g_2 = 1.5$ mm, $t = 0.3$ mm.*

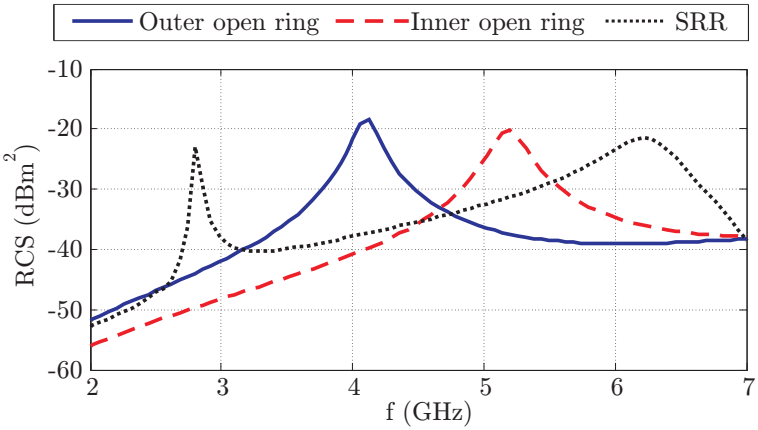


Figure 2.35: *Computed monostatic radar cross section in free space for the SRR of figure 2.34. Solid line: stand alone outer open ring. Dashed line: stand alone inner open ring. Dotted line: SRR.*

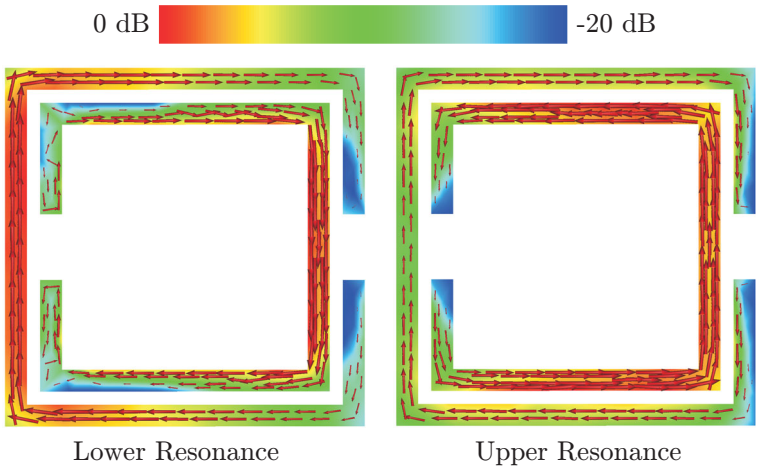


Figure 2.36: *Simulated current distributions. Each image is individually normalized and displayed on a 20 dB scale. Arrows are scaled according to the normalized current levels.*

Antenna design

The SRR loaded monopole antenna geometry consists of a monopole loaded with two identical SRRs (figure 2.37). The triple band performance could be achieved with a single SRR but a symmetric structure will result in highly symmetric radiation patterns with low crosspolar levels. The main drawback of the symmetric structure is an increase in the antenna width.

The design begins by considering the middle band (3.244 - 4.742 GHz) and adjusting the monopole parameters L_m, w_m, L_0 and w_0 to obtain impedance matching in a large

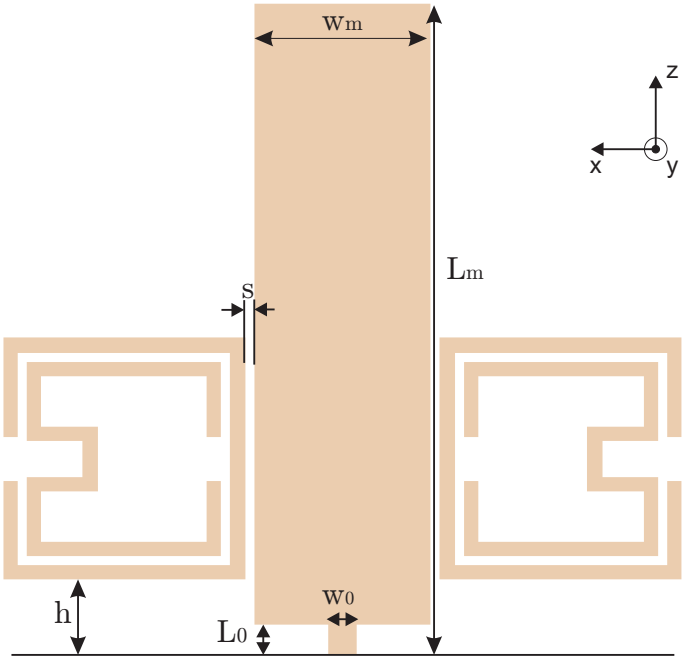


Figure 2.37: *SRR loaded monopole. The structure was fabricated on Rogers RO4003C 0.508 mm substrate (ε_r = 3.38, $\tan\delta$ = 0.0027). L_m = 22 mm, w_m = 6 mm, L_0 = 1 mm, w_0 = 1 mm, h = 2.5 mm, s = 0.3 mm. SRRs dimensions are the same as in figure 2.34.*

bandwidth centered just below the limit of this band, because according to figure 2.38, the presence of the SRRs will introduce an upward frequency shift in the monopole matching band. Using the values of figure 2.37, about 1 GHz of bandwidth centered at 3 GHz is obtained.

The SRRs are designed to resonate just above the upper limits of the 802.11 standard lower (2.4 - 2.5 GHz) and upper (4.9 - 5.875 GHz) bands. The SRR shown in figure 2.34, which resonates at approximately 2.8 GHz and 6.2 GHz (figure 2.35), is used. The orientation of the SRRs with respect to the monopole is selected in order to obtain maximum coupling. Other orientations do not provide enough coupling to obtain good matching levels over all the operating bands.

Figure 2.38 shows that when loading the monopole with the SRRs, two peaks appear at the SRRs resonances in both S_{11} and input impedance. Just before these peaks, two minimums in S_{11} appear used to allocate the lower and upper bands of the universal 802.11 standard. The mid-band response shifts higher in frequency when the SRRs are introduced.

The return losses and the position of the minimums can be slightly adjusted with the

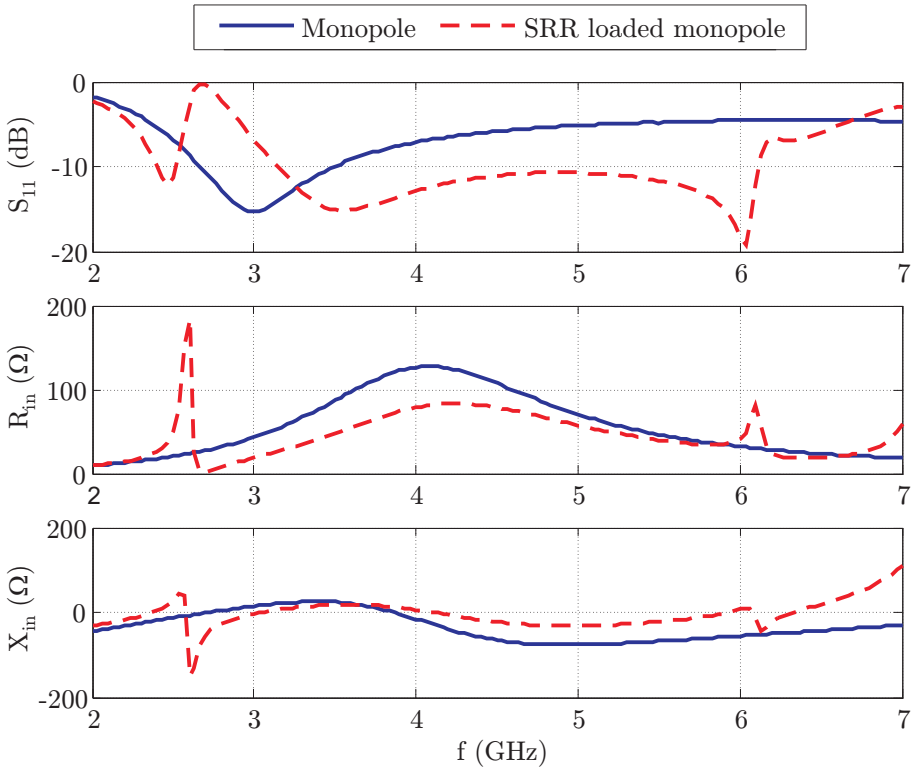


Figure 2.38: *Simulations comparing the stand alone monopole and the SRR loaded monopole.*

parameters s and h defined in 2.37. The coupling between SRRs and monopole reduces with increased separation s, resulting in smaller peaks in the impedance and less shift in the middle band. The SRRs to ground plane separation h sets the SRRs load position on the monopole, with larger values of h having a less significant loading effect on the input impedance. A fine tuning of these parameters gives the values of figure 2.37 and the results of figure 2.38 that meet the requirements.

The simulated antenna radiation efficiency is plotted in figure 2.39. The efficiency minimums introduced by the resonators losses are out of band and coincide with the S_{11} peaks, as a result the radiation efficiency in the operating bands is not expected to be significantly degraded by the high losses of the resonators at their resonance frequencies.

Prototype antenna: Fabrication and measurement

Figure 2.40(a) shows the fabricated prototype mounted on a large ground plane. The measured S_{11} parameters agree well with simulations (figure 2.40(b)). However the upper SRRs resonance is slightly shifted in frequency and its effects on the monopole input impedance are weaker than the simulation predictions. This may be due to tolerances

in the fabrication process. Anyway, according to table 2.5 a clear matching band can be observed at 2.45 GHz, and a broadband matching frequency range is obtained from 3 GHz to 6.2 GHz, covering the requirements of the ISM bands and the 802.11 and 802.15 standards.

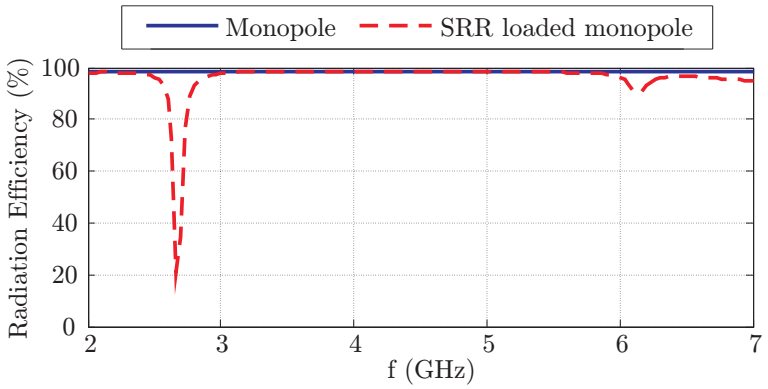


Figure 2.39: *Simulated radiation efficiency of the stand alone monopole and the SRR loaded monopole.*

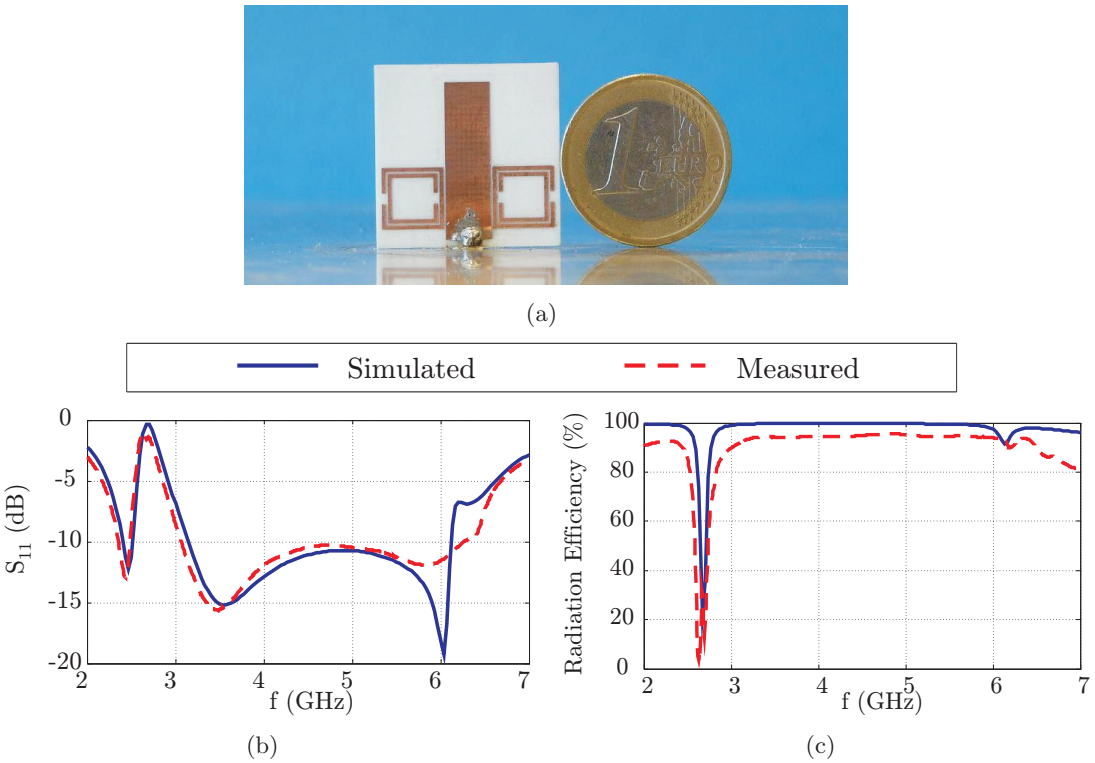


Figure 2.40: *SRR loaded monopole prototype antenna. (a) Prototype antenna mounted on a large ground plane. (b) Simulated and measured S_{11} parameters. (c) Simulated and measured radiation efficiency.*

Table 2.5: *Measured operating bands of the triple band SRR loaded monopole antenna (figure 2.40(a)). The second and the third bands have been combined obtaining a broadband matching frequency range from 3 GHz to 6.2 GHz.*

SRR loaded monopole antenna		
Operating band (GHz)	Center frequency (GHz)	Relative bandwidth(%)
2.35 - 2.5	2.43	6.2
3.06 - 6.25	4.65	68.5

The radiation efficiency (figure 2.40(c)) was measured using the method presented in [69]. By virtue of the off-resonance loading principle, the in-band radiation efficiency is practically unaffected by the SRRs losses, resulting in efficiency levels higher than 85% over all the operating bands. These measurements confirm that, although FEKO radiation efficiency simulation is optimistic, it predicts the overall behavior of the structure and the efficiency minimums position. At frequencies lower than 2 GHz and frequencies higher than 6.4 GHz efficiency measurements are distorted due to the limitations of the measurement method [69].

Figure 2.41 shows the measured copolar and crosspolar components of the normalized radiation pattern cuts in the main planes ($\phi = 0°$, $\phi = 90°$ and $\theta = 90°$) at 2.45 GHz, 3.5 GHz and 5.8 GHz. The E-plane cuts are quite similar for all the frequencies and the H-plane cuts are essentially omnidirectional. Furthermore, due to the antenna symmetry, measured crosspolar levels are low, normally less than -20 dB. The measured gain values range from 5 dB to 6 dB over all the bands, and are in agreement with the measured efficiency values (figure 2.40(c)).

These results demonstrate that the antenna presented in this section offers good triple band performance and could be used for the central unit, taking into account that the central unit could also use the 3.2 - 4.7 GHz frequency band of the 802.15 standard to transmit information.

2.3.4 Triple band resonator loaded monopole antenna based on the use of an asymmetric structure

As has been mentioned with fractal geometries, asymmetric structures are a good option to increase design flexibility in multiband antennas. In this section a novel antenna design that uses two different loading resonators to obtain a triple band asymmetric monopole will be shown. The use of an asymmetric structure will introduce a new degree of freedom in the design process and with the off-resonance loading principle the resonators losses in

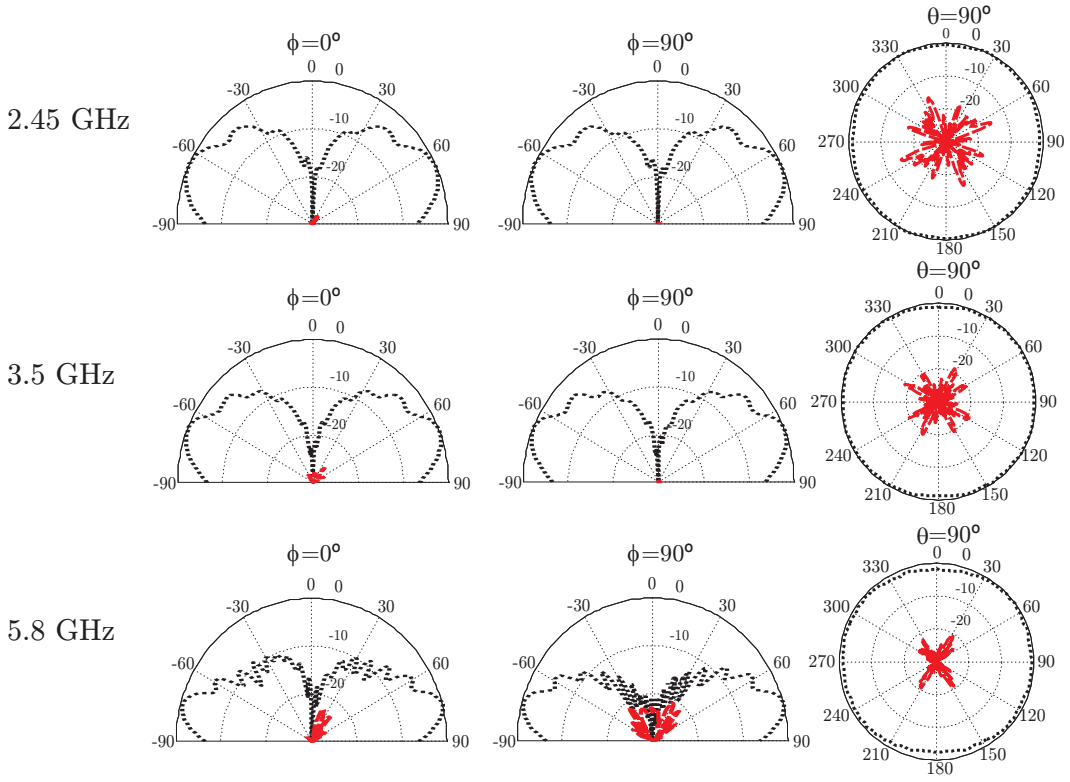


Figure 2.41: *Measured normalized radiation patterns of the SRR loaded monopole. Each cut has been normalized with reference to its maximum. Black dotted lines are used for the copolar component; red dashed lines are used for the crosspolar component.*

the matching bands will be avoided.

Antenna design

The antenna geometry consists of an asymmetric structure based on a monopole loaded with two resonant particles with different resonance frequency (figure 2.42). Figure 2.43 shows the existence of three matching bands around 2.45 GHz, 3.5 GHz and 5.4 GHz.

The first S_{11} minimum is used to cover the lower band of the 802.11 standard (b / g) and the 2.45 GHz ISM band. As can be seen in figure 2.44 currents have the typical distribution of a $\lambda/4$ monopole with significant currents in the resonators, especially in the largest one. Therefore this band corresponds to the first resonance of the straight monopole, taking into account that it is slightly frequency-shifted due to the resonator loading effects.

With the second S_{11} minimum, the 3.5 GHz band of WiMAX can be covered. This minimum basically appears due to the large resonator effects (it resonates at a frequency

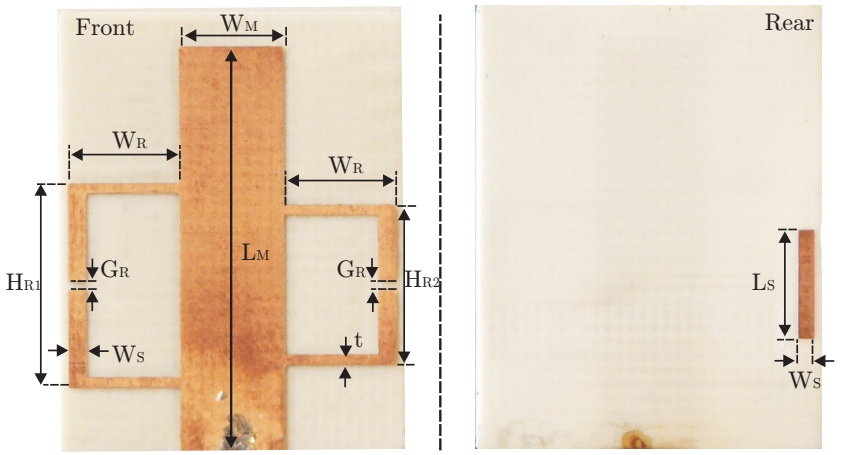


Figure 2.42: *Fabricated prototype. In the rear side a conductor strip is used to reduce the resonance frequency of the large resonator.*

around 2.9 GHz) and its position can be controlled adjusting the resonance frequency of this resonator. In the design of this resonator, a conductor strip in the rear side of the substrate is used to increase its capacitance and decrease the resonance frequency without a significant increase in the resonator dimensions. In this frequency band currents in the small resonator are also significant (figure 2.44), and hence it will be slightly frequency-shifted due to the effects of this resonator.

Finally, the upper band of the international 802.11 standard (a / h / j) and the 5.8 GHz ISM band are covered by combining the third and fourth S_{11} minimums. According to figure 2.44, the third S_{11} minimum is mainly due to the minor resonator effects (it resonates at a frequency around 4.5 GHz) and its position can be controlled adjusting the dimensions of this resonator. The fourth S_{11} minimum is due to the monopole second resonance. However, as in the loaded monopole of section 2.3.2, at this frequency currents tend to follow the perimeter of the structure. Therefore, this S_{11} minimum is affected by the additional path introduced by the resonator arms. For this reason, the position of this minimum can be controlled adjusting the resonators width W_R. Tuning the separation between both S_{11} minimums the available bandwidth of this band can be optimized, taking into account that a clear trade-off exists between the available bandwidth and the matching levels at the center of the band.

Figure 2.43 shows a parametric analysis of the variations of S_{11} parameter with respect to the most important design parameters of figure 2.42. The monopole length modifies the position of all the matching bands (figure 2.43(a)), while the monopole width does not significantly change the position of the bands, although it affects the overall matching levels (figure 2.43(b)).

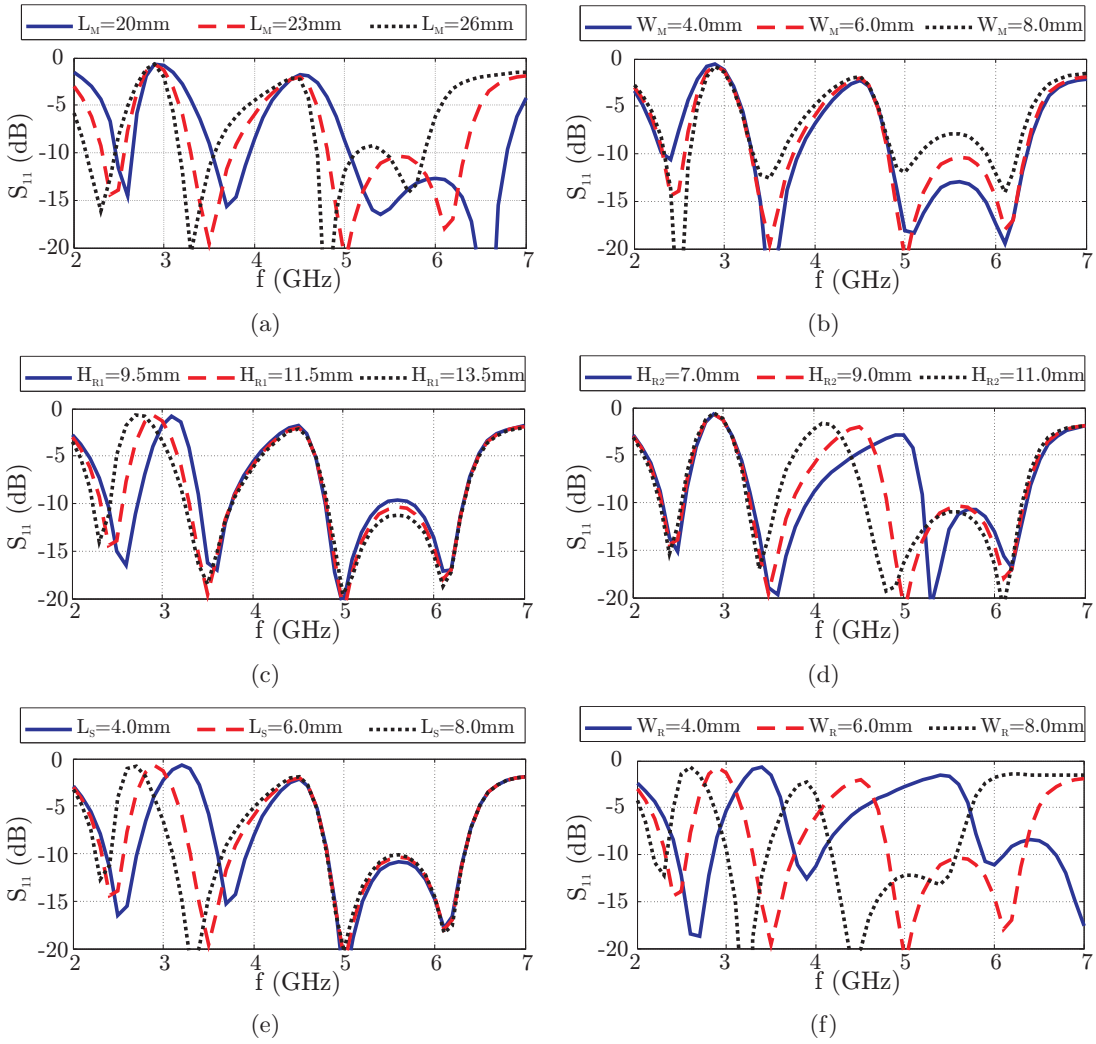


Figure 2.43: *Parametric study of the variation of the S_{11} parameter with respect to (a) the length of the monopole L_M, (b) the width of the monopole W_M, (c) (d) the height of the resonators H_{R1}, H_{R2}, (e) the length of the rear side strip L_S and (f) the width of the resonators W_R.*

The resonance frequency of the large resonator lies between the first and the second S_{11} minimums: for this reason when this resonator is modified, these two minimums will be affected. This is proved in figure 2.43(c) and figure 2.43(e), where the resonance frequency is changed by modifying the inductance and the capacitance respectively. The resonance frequency of the small resonator lies between the second and the third S_{11} minimums, therefore the modifications applied to this resonator will affect the properties of both minimums(figures 2.43(d)).

Finally, when the resonator width is changed all the S_{11} minimums are moved (figure 2.43(f)). This is logical, since this width will change the resonance frequency of both

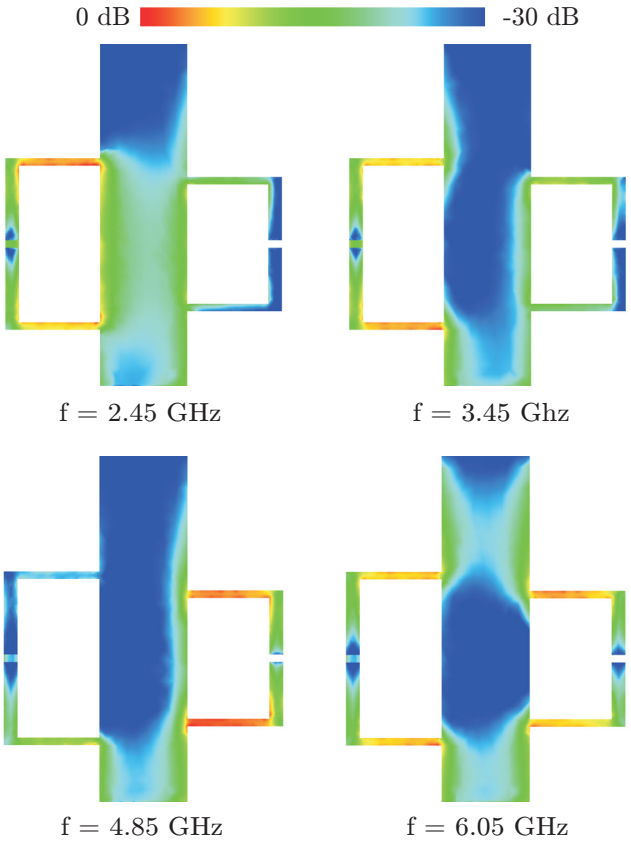


Figure 2.44: *Simulated current distributions at the S_{11} minimums. Each image is individually normalized and displayed on a 30 dB scale.*

resonators, and as it has been shown before, this will change the position of the first three S_{11} minimums. With reference to the fourth S_{11} minimum, the additional path added by the resonators arms will also change, and as a result, this minimum will be frequency shifted.

Prototype antenna: Fabrication and measurement

Taking into account the aforementioned trends, a monopole antenna for ISM / WLAN / WiMAX services was designed. The whole monopole dimensions are 20 mm x 25 mm and it was fabricated on Rogers RO4003C substrate ($\epsilon_r = 3.38$, $tan\delta = 0.0021$). Table 2.6 shows the dimensions of the prototype antenna.

In figure 2.45 three matching bands at 2.45 GHz, 3.5 GHz and 5.4 GHz are observed, with relative bandwidths of 14.7 %, 18.9 % and 23.2 % respectively, enough to cover the ISM, WLAN (802.11 a/b/g/h/j) and WiMAX requirements. More detailed information about the operating bands can be found in table 2.7.

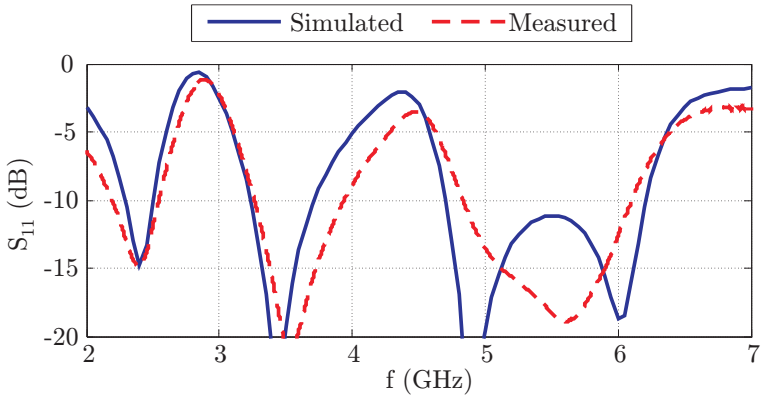


Figure 2.45: *Measured and simulated S_{11} parameter of the asymmetric loaded monopole antenna. Blue solid line shows simulated results, and red dashed line shows measured results.*

Table 2.6: *Dimensions of the asymmetric prototype antenna.*

L_M	W_M	L_S	W_S	G_R
23 mm	5.8 mm	6.1 mm	0.9 mm	0.5 mm

t	H_{R1}	H_{R2}	W_R
0.5 mm	11.5 mm	9 mm	6.3 mm

Table 2.7: *Measured operating bands of the triple band asymmetric resonator loaded monopole antenna (figure 2.42).*

Asymmetric resonator loaded monopole antenna		
Operating band (GHz)	Center frequency (GHz)	Relative bandwidth(%)
2.2 - 2.55	2.375	14.7
3.26 - 3.94	3.6	18.9
4.83 - 6.10	5.46	23.2

The measured and simulated position of the efficiency minimums agree well (figure 2.46). These efficiency minimums are related to the resonance of the resonators and again, thanks to the use of the off-resonance loading principle, they are out of the antenna operating bands, and therefore do not degrade the in-band total efficiency. The measured total efficiency is higher than 80 % in all the three bands.

The measured radiation patterns of the proposed design are plotted in figure 2.47. E-plane radiation patterns are similar in all bands and have the typical shape expected from a quarter wavelength monopole. H plane radiation pattern cuts are essentially omnidirectional in all the bands. Crosspolar levels will be slightly degraded due to the antenna asymmetry, however in the radiation pattern cuts shown in figure 2.47, crosspolar

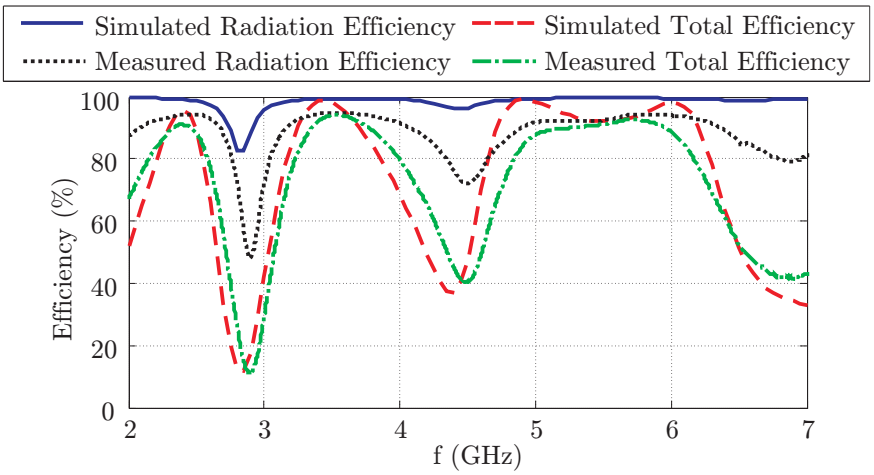


Figure 2.46: *Measured and simulated results for the radiation efficiency and the total efficiency. Total efficiency takes into account the mismatch loss.*

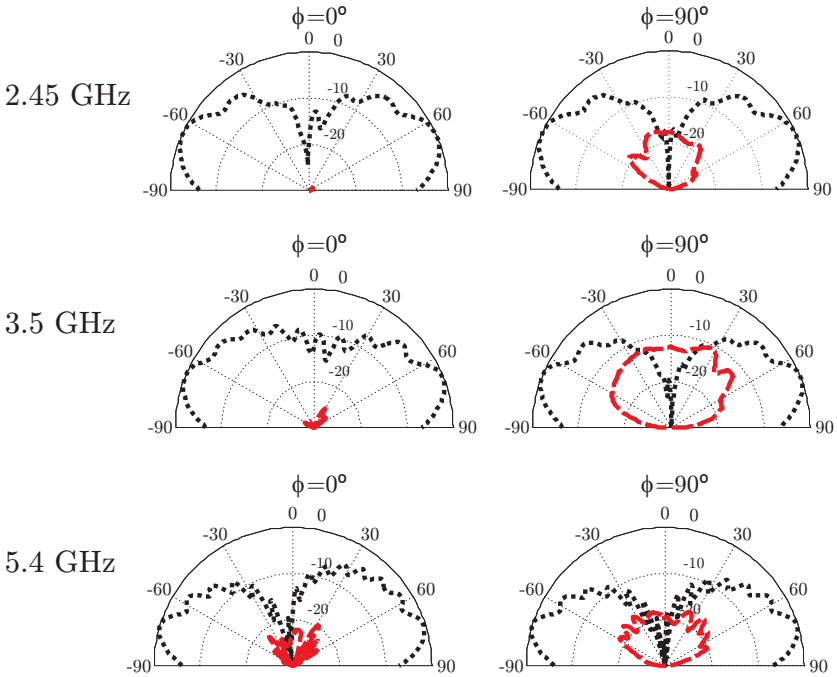


Figure 2.47: *Measured radiation patterns of the asymmetric loaded monopole in the three operating bands. Each cut has been normalized with reference to its maximum. Black dotted lines are used for the copolar component; red dashed lines are used for the crosspolar component.*

levels are below -15 dB in the directions of maximum radiation. The measured gain values range from 5 dB to 6 dB over all the bands.

Measurements and simulations show that the proposed antenna offers good matching levels, measured total efficiency levels of over 80 % and similar radiation patterns along all the three bands, therefore it is also suitable for the central unit. Furthermore, with this novel antenna the central unit capabilities could be improved using a WiMAX connection to send the collected information to the Internet without the need for any additional radiating element.

2.4 Chapter conclusions

In the central unit, multiband radiators are desirable to be able to receive information from sensors using the 2.45 GHz and 5.8 GHz ISM bands, and simultaneously send this information to the Internet using the 2.45 GHz and 5 GHz WLAN bands with a single radiating element. The solutions presented in this chapter are based on:

- Fractal shapes.
- Structures loaded with resonators.

The most commonly used fractal shape for the design of multiband antennas is the Sierpinski gasket due to its self similarity property at different scale levels. Fractal shapes have usually been utilized in the monopole element assuming that it is on a large ground plane, however they can be also applied to the ground plane. As an example a fractal-based ground plane inspired by the Sierpinski gasket has been presented. The modified ground plane adds new resonances that can be combined with the monopole element resonances to improve the antenna multiband performance. As proof of concept, two novel antenna designs have been built and measured.

The first one is a dual band antenna based on a triangular monopole mounted on a fractal ground plane. The novel ground plane has allowed us to add new operating bands to the triangular monopole without degrading radiation patterns and efficiency. Through current distribution analysis, it has been shown that currents in the ground plane scale with frequency, and thanks to the self-similarity property of fractal shapes, the current distributions in the different frequency bands are the same but at different scale levels. This is the main reason why this kind of ground plane offers good log-periodic multiband performance.

A Sierpinski monopole on a fractal-based ground plane has been also presented. The use of the fractal shape in both elements gives more flexibility to the design process. When the Sierpinski gaskets used in both elements are equal, the ground plane reinforces the monopole resonances, while if they are slightly different, resonance frequencies are not the same and the available bandwidths of the whole structure can be increased. Adjusting the height of the monopole above the ground plane, a fine tuning process can be carried out and the overall antenna performance can be improved. Comparing the first antenna design with this second, the combination of the fractal shape in both elements improves the available bandwidth in the 2.45 GHz band. These two multiband antennas cover the 2.45 GHz and 5.8 GHz ISM bands, and the 2.45 GHz and 5 GHz WLAN bands, and hence they can be perfectly used as radiating element for the sensor network central unit.

The second solution proposed in this chapter to obtain multiband behavior is resonator loaded monopoles. In order to improve the antenna performance it is advisable to use the off-resonance loading principle. This means that the resonators are designed to allocate their resonances out of the antenna operating bands. With this technique new matching bands are introduced without significantly degrading the in-band radiation efficiency.

Based on this principle a simple design procedure for a dual-band ISM / WLAN monopole antenna based on capacitively loaded magnetic resonators has been presented. A fabricated prototype that easily meets the specifications was demonstrated, showing good matching, high efficiency levels and similar radiation patterns in both bands. This prototype antenna covers the ISM and WLAN bands, and hence it is also suitable for the central unit.

The same concept can be also applied to introduce more operating bands. As an example, two novel triple band antennas have been presented. In the first, the triple band performance is achieved using dual band SRRs. The lower and upper bands positions are essentially controlled by the SRRs dual band behavior whereas the monopole dimensions define the middle band. A prototype that covers the universal 802.11 standard, the 5.8 GHz ISM bands and one band of the 802.15 standard (3.24 - 4.74 GHz) with radiation efficiencies higher than 85% was fabricated and measured.

In the second example, the triple band performance is achieved using simple resonators in an asymmetric structure. The asymmetric structure allows for the addition of new operating bands, and the position of these bands can be easily controlled by the design parameters of the loading resonators. An asymmetric prototype antenna for the ISM, 802.11 standard and 3.5 GHz WiMAX bands was fabricated and measured, proving the good performance offered by this topology. Moreover, the effects of two different dual band

resonators could be combined in an asymmetric structure to further improve the antenna multiband behavior.

These two antennas cover the ISM and WLAN bands, and therefore are suitable for the central unit. Furthermore the third band could be used to improve the system performance with a WiMAX connection for those places away from any WLAN access point.

In this chapter several novel antenna designs suitable for the central unit of the sensor network have been presented. In a real application, if we can define the ground plane geometry, the solutions based on fractal ground planes can be applied. With these solutions, the ground plane area could be covered in a transparent plastic, and the antenna could even be seen as a decorative element. However, if access to the ground plane design is not available, the solutions based on resonator loaded monopoles must be used. Both solutions could be also combined to improve performance.

Chapter 3

Small silicon integrated antennas for the remote sensors

This chapter of the thesis will focus on the design of antennas for the remote sensors. As has been said in the introduction, sensors dimensions and cost are key factors to ensure system viability, therefore small and cheap sensors are desirable. One possible solution to achieve these requirements is the use of integrated silicon antennas fabricated on standard CMOS technology silicon substrates, where the antenna, the electronic circuits and the sensor are made together in the same chip, obtaining a system-on-chip solution.

Firstly the performance offered by different miniaturization techniques when they are applied to antennas built on a low resistivity substrate will be analyzed. Next, micromachining techniques will be presented as a solution to improve antenna efficiency, and finally the performance degradation due to the presence of parasitic elements will be studied.

3.1 Introduction to silicon integrated antennas

Packaging and assembly of radio front-ends, especially at high frequencies, is currently a significant cost of the total product. Hence, monolithic integration of all the high frequency components on a single semiconductor chip has the potential of lowering the price of transceiver units to acceptable levels for mass market applications [74]. The fabrication of the antenna and the electronic circuits in the same chip reduces assembly costs and the system dimensions [75]. Despite their potential cost and integration benefits, the use of on-chip antennas has not yet become widespread. One of the main reasons is that the antenna is one of the most difficult RF-components to integrate on-chip with active circuits.

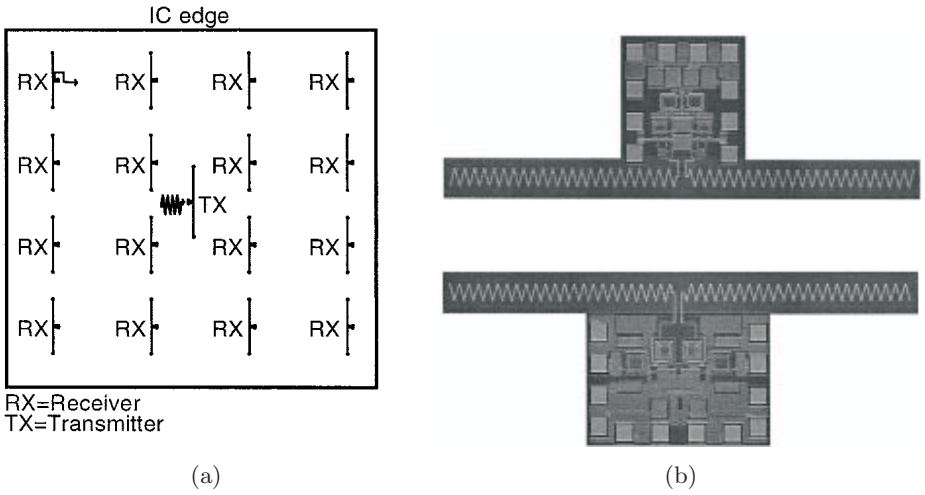


Figure 3.1: *Example of intra-chip communication system. (a) Conceptual system illustration of intra-chip wireless interconnected system for clock signal distribution. (b) 0.18 μm clock transmitter and receiver with zigzag dipole antennas [78].*

In order to reduce fabrication costs and obtain real system on-chip devices, it is also desirable to make the antenna fabrication process compatible with standard CMOS technologies using a silicon substrate not optimized for antennas production. The low bulk resistivity silicon of the wafers normally used in standard CMOS processes for the fabrication of active devices decreases the antenna radiation efficiency drastically. Furthermore this efficiency degradation is emphasized when miniaturization techniques are applied to reduce antenna dimensions [76]. All these problems mean that integrated antennas design is not an easy process.

In literature, several examples of silicon integrated antennas for intra-chip and inter-chip communications can be found. As an example of intra-chip communications, in [77, 78] integrated antennas are used for the distribution of the clock signal to minimize the delay (figure 3.1). In a similar way, integrated antennas could be also used to transmit the clock signal among different chips.

Some of the first examples of antennas for inter-chip communications and for communications with an external receiver were presented in [79], where Buechler showed some antennas built on silicon substrates (figure 3.2(a)). In [80], the folded slot of figure 3.2(b) is proposed as basic radiating element for automotive short range radar applications in the 79 GHz band.

At the millimeter wave frequency range wavelength is small and therefore complex antennas can be integrated in a reduced area. In [81] a 1.3 mm quasi-Yagi antenna with one driver, two directors and a truncated ground plane that acts as a reflector is presented

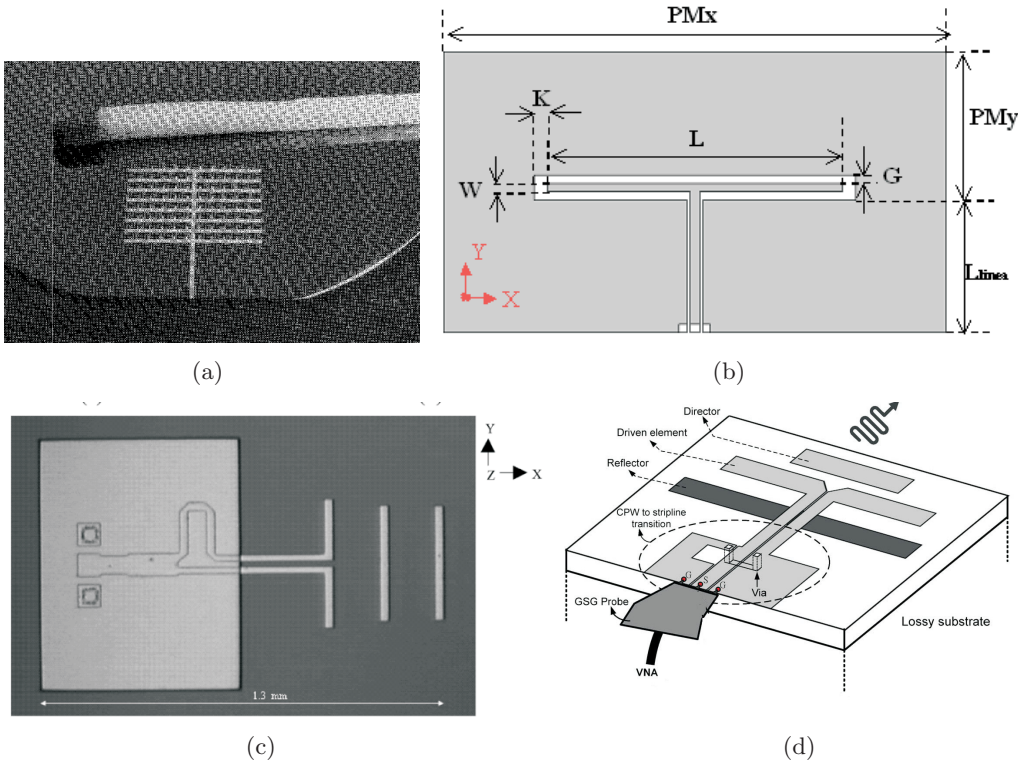


Figure 3.2: *Examples of integrated antennas operating at frequencies higher than 60 GHz. (a) Array antenna fabricated on high resistivity silicon presented by Buechler in 1986 [79]. (b) Folded slot antenna designed at 79 GHz [80]. (c) and (d) 60 GHz on-chip Yagi antennas [81, 82].*

(figure 3.2(c)). Another example can be found in [82], where a director element, a driven element and a reflector element are used on the design of a Yagi antenna with dimensions of 1.1 mm x 0.95 mm (figure 3.2(d)).

The aforementioned integrated antenna examples work in the frequency bands of 60 GHz or higher. In [75, 83, 84] Lin studies the performance of integrated antennas in the 24 GHz ISM band. These antennas are fabricated using the conductor and dielectric layers available in mainstream silicon technologies to keep the integration of on-chip antennas viable and inexpensive [75].

Figure 3.3 shows the layout of the zigzag dipole antenna presented by Lin in [83], a schematic representation of the measurement setup and the measured antenna pair gain for different environments. The antenna pair gain includes the gain of the transmitting and receiving antennas and the propagation losses. At a distance of 1 m, antenna pair gains of less than -80 dB have been measured. In spite of these low values, it can be seen that the zigzag dipole increases the antenna pair gain by 35 dB with respect to the situation

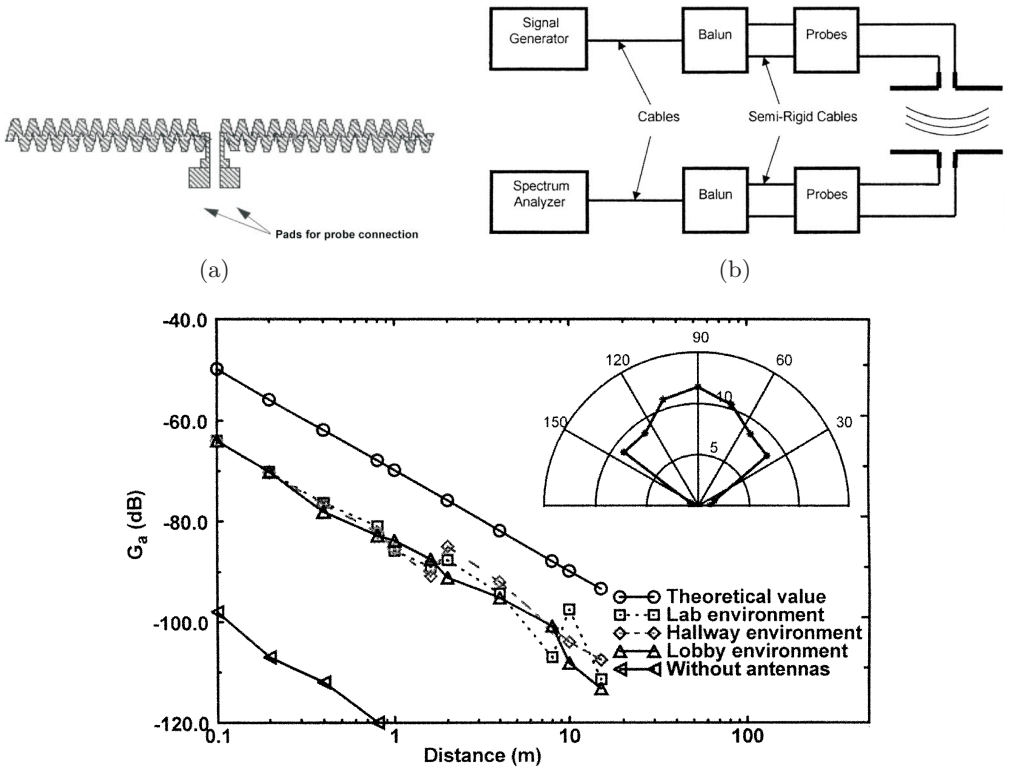
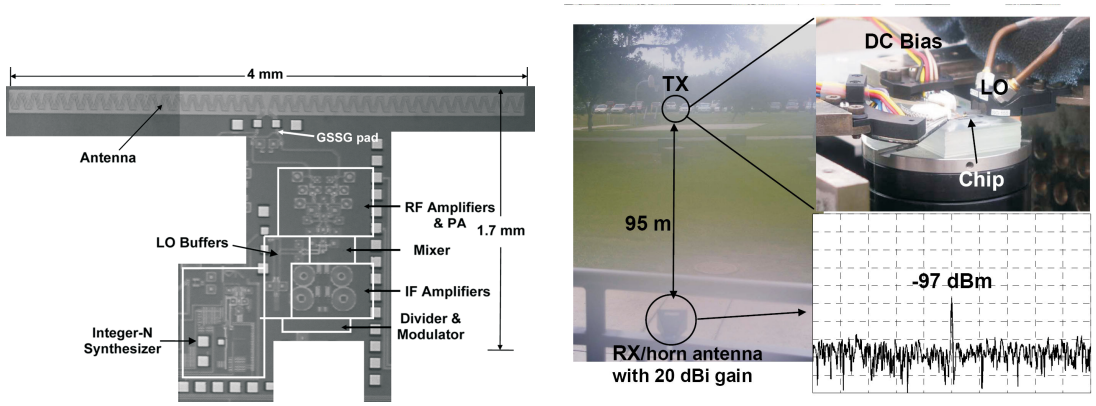


Figure 3.3: *Integrated antenna presented by Lin in [83]. (a) Layout of the 2 mm zigzag dipole. (b) Antenna measurement setup with a signal generator and a spectrum analyzer. (c) Antenna pair gain as a function of the distance in three different environments at 24 GHz. The inset shows the 2 mm zigzag dipole antenna radiation pattern.*

where the probes are the only radiating elements.

Other examples of silicon integrated antennas operating at frequencies higher than 10 GHz can be found in [85–89], where several techniques and topologies, such as IFA structures, micromachining or dielectric resonators, are used in order to improve the antenna performance. The previous references are focused on the antenna design, but there are also some references where the whole system is analyzed. For instance in [90, 91] a fully integrated transmitter with an on-chip dipole antenna is presented, showing that complete system integration with high system ranges is possible (figure 3.4).

In some applications, such as in implantable devices, frequencies lower than 10 GHz are desirable to reduce the additional propagation losses introduced by body tissues [92, 93]. At lower frequency bands the radiating element dimensions become a problem because antennas are larger, therefore miniaturization techniques must be applied [76] to



(a) (b)

Figure 3.4: *Fully integrated transmitter with an on-chip dipole antenna at 24 GHz [91]. (a) Photomicrograph of the fully integrated transmitter and frequency synthesizer. (b) Reception of the signal generated by a transmitter IC with an on-chip antenna using a 20 dBi gain horn antenna.*

allow antenna integration without an excessive enlargement of the chip dimensions. Some examples of integrated antennas operating at frequencies lower than 10 GHz can be found in [75, 76, 94]. For instance, in [75] Lin presented the gain plot of figure 3.5, showing that the antenna pair gain at 5.8 GHZ is significantly better than at 24 GHz.

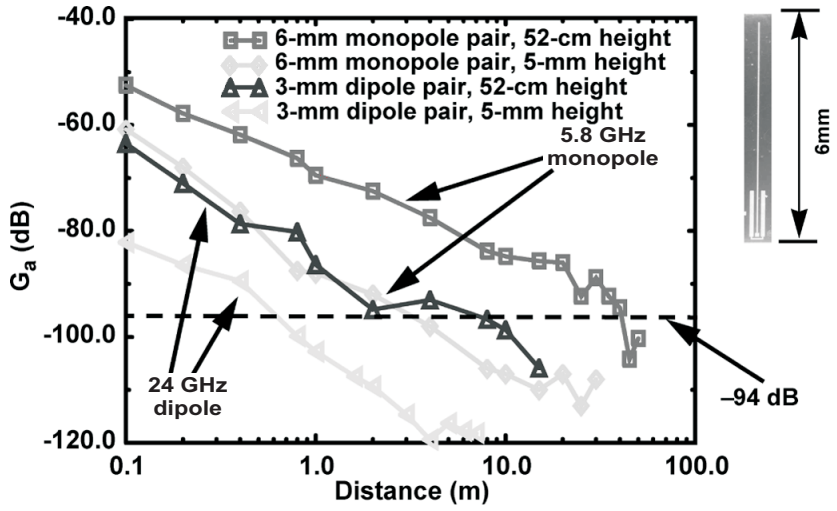


Figure 3.5: *Antenna pair gain as a function of the distance of a 5.8 GHz monopole and a 24 GHz dipole antenna. The antennas are located 52 cm and 5 mm above the ground [75].*

3.2 Prototype antennas fabrication and measurement

The objective of this chapter is the design, fabrication and measurement of standard CMOS silicon integrated antennas operating in the 5.8 GHz ISM band. A thorough study of the fabrication process is not included among the objectives of this thesis. Nevertheless some basic knowledge of the fabrication technology is necessary due to the fact that the resulting dielectric/metal layers structure and the materials properties will critically affect antenna performance and the design process.

In order to make integration of an on-chip antenna viable, the fabrication process must be inexpensive. This means that the antenna must be fabricated using the conductor and dielectric layers available in standard CMOS technologies and a silicon substrate compatible with these technologies.

The prototype antennas shown in this thesis were made in a dedicated process in the clean room of the Centro Nacional de Microelectronica - Instituto de Microelectronica de Barcelona (CNM-IMB, [95]). The process was intended to yield a metal/dielectric structure equivalent to a standard 0.18 μm RF CMOS, that is, a top metal of 2 μm thick Aluminum (Al) over 6 μm of SiO_2 (figure 3.6). The substrate was low resistivity silicon with a theoretical resistivity of 20 Ohm·cm.

Briefly, the main steps of the fabrication process were:

1. Start with a 4" wafer of P doped silicon with a resistivity of 20 Ohm·cm.
2. Growth of 2 μm of SiO_2.
3. Additional 4 μm-thick deposition of SiO_2.
4. 2 μm thick deposition of Al.
5. Al patterning with standard photolithography and etching process.

For the antennas fabricated during this thesis, two chip sizes were used. Firstly, the chip dimensions were 12.5 mm x 2 mm. Then, they were reduced to 5.25 mm x 1.9 mm to obtain smaller devices. Since the available area is very small, antenna miniaturization techniques will be necessary.

According to figure 3.6, the antenna is at a distance of 6 μm from the silicon substrate, and hence the silicon effects on the antenna performance will be important. The relative permittivity of silicon is high, $\epsilon_r = 11.9$, so it reduces the antenna dimensions. The problem

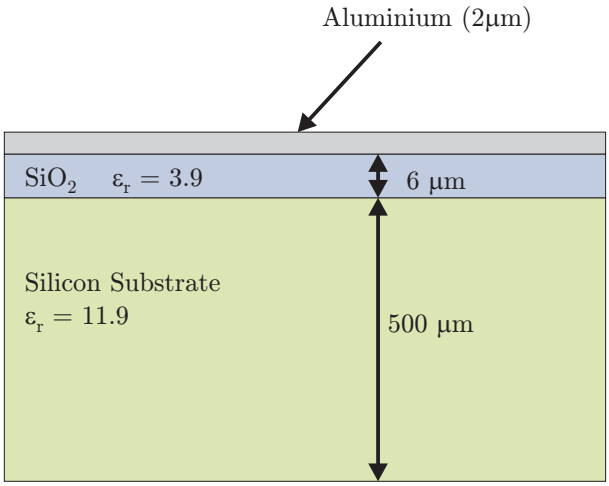


Figure 3.6: *Simplified metal/dielectric layer structure of the fabricated prototypes. It is equivalent to a standard 0.18 μm RF CMOS technology.*

is that the low resistivity of silicon significantly degrades the antenna radiation efficiency and gain.

The SiO_2 layer helps to isolate the antenna from the silicon substrate, slightly improving antenna gain. A comparison of the simulated antenna input parameters and radiation efficiency with and without the SiO_2 layer can be found in figure 3.7, showing as in the case of using the SiO_2 layer losses are slightly lower and the impedance curves are less smooth.

Numerical simulations must take into account the dielectric layers structure of the fabricated prototypes (figure 3.6). When infinite dielectrics are considered, simulations can be carried out without problems. However, some simulation problems arise with the Method of Moments (MoM) code of FEKO software when the finite size of very thin dielectric layers is taken into account. The meshing process of a thin layer is difficult because the number of triangular mesh elements is large and with a high aspect ratio. This increases the computational requirements (time and memory) and can introduce numerical errors that generate incorrect simulation results.

Finite dielectrics are necessary since the chip size is an important parameter that affects the antenna performance. To support this affirmation, the same antenna design was fabricated on three chips with different widths (2 mm, 4 mm and 6 mm). Frequency shifts higher than 500 MHz were observed due to the variation of the effective relative permittivity.

One option to avoid these problems is the use of a simulation model without the

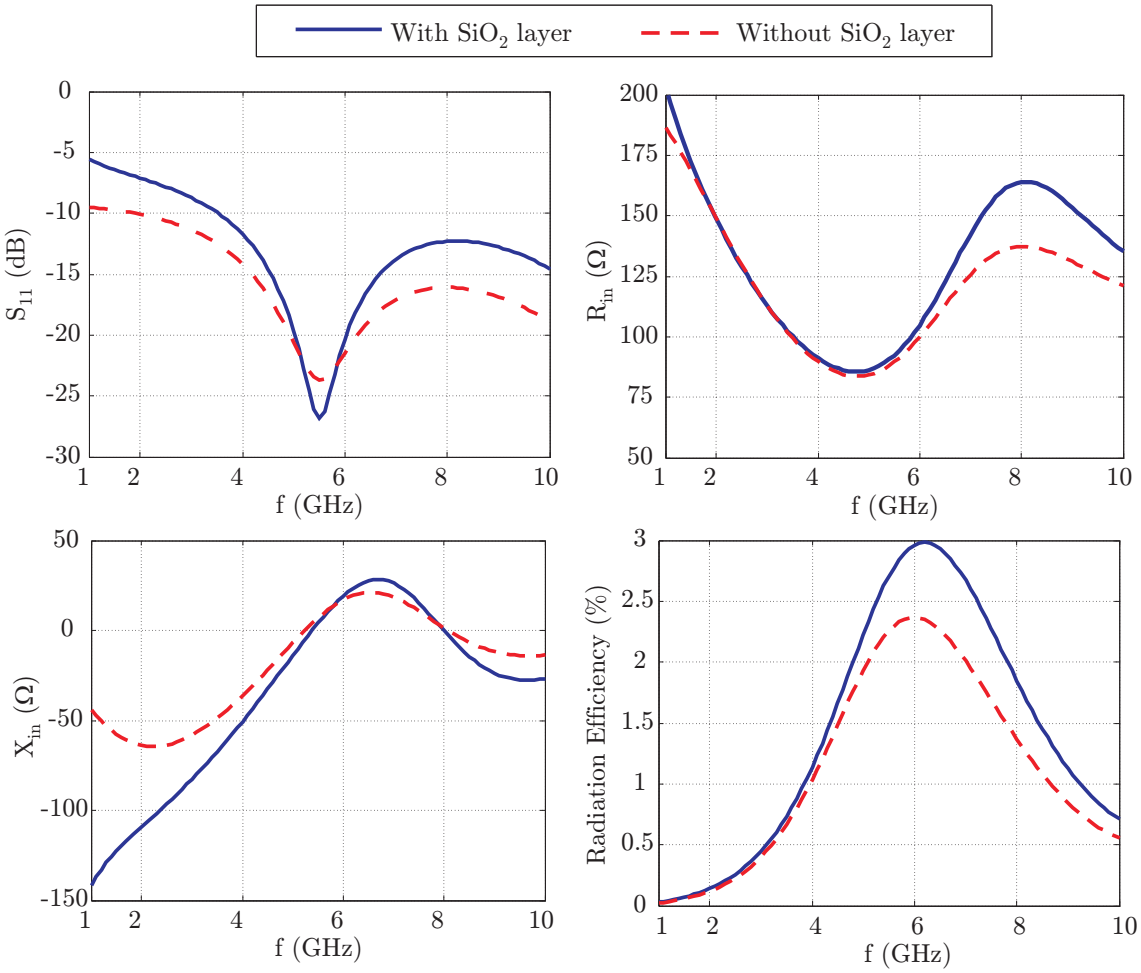


Figure 3.7: *Antenna input parameters and radiation efficiency with and without the SiO_2 layer. The silicon conductivity is $\sigma = 5$ S/m. The antenna is a meandered dipole and the chip dimensions are 5.25 mm x 1.9 mm.*

SiO_2 layer. Due to the thinness of this layer, the results would be quite similar to the real ones (figure 3.7). Nevertheless, if we want to include the SiO_2 effects in the simulations, there are other numerical methods (FDTD or FEM) that are more suitable for the study of complex dielectric structures. In our case we decided to use the hybrid Method of Moments − Finite Elements Method (MOM-FEM) FEKO code [70].

A precise simulation model was necessary in order to achieve good agreement between simulations and measurements, for this reason, the parameters of the wafers used in the antennas fabrication process were measured. Wafers thickness measurements gave the same value as the theoretical one (500 μm). The SiO_2 layer thickness ranged from 5.8 μm to 6.7 μm, and hence simulations were made using the theoretical value of 6 μm. Important differences were observed in the conductivity ranging from 11.5 Ω·cm to 13.4

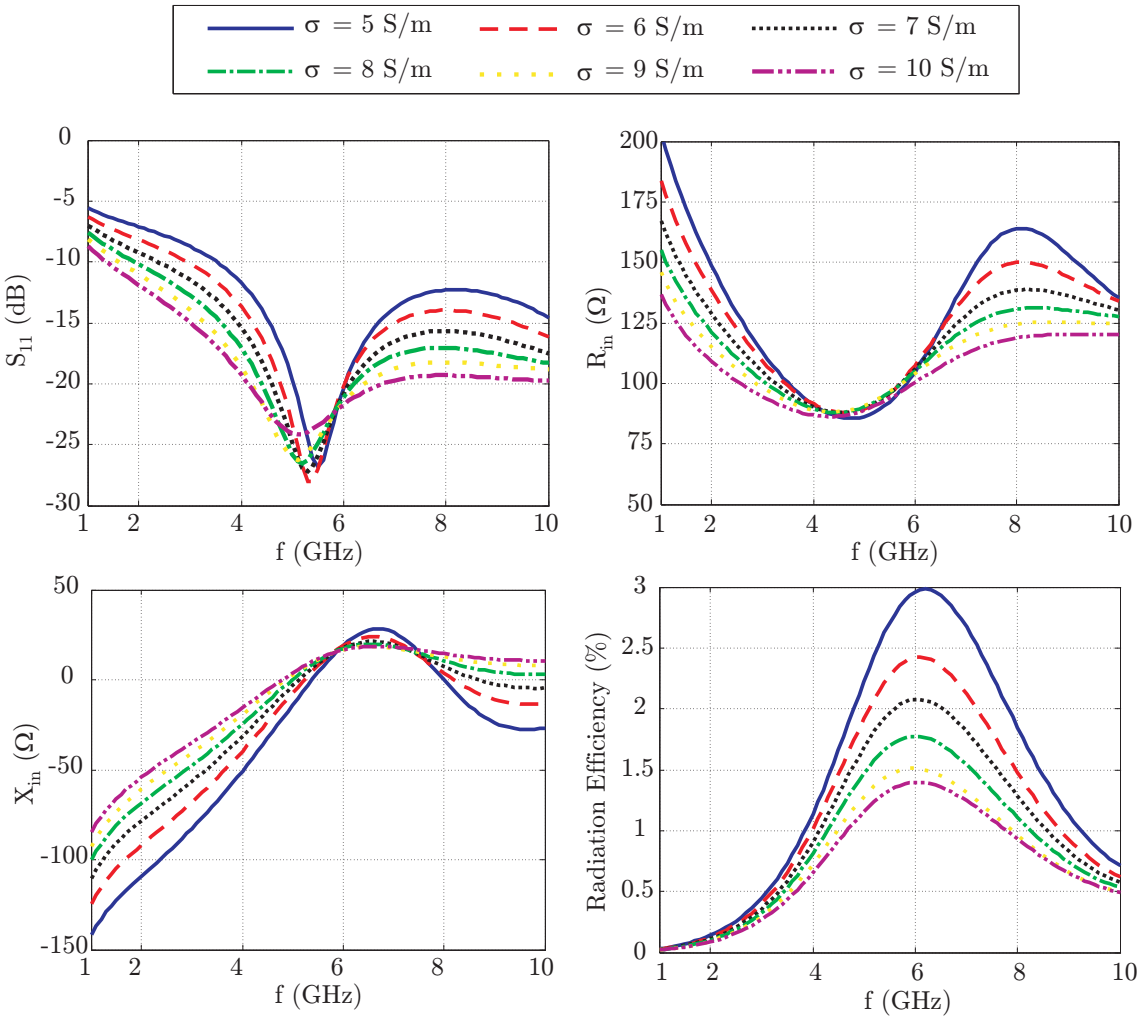
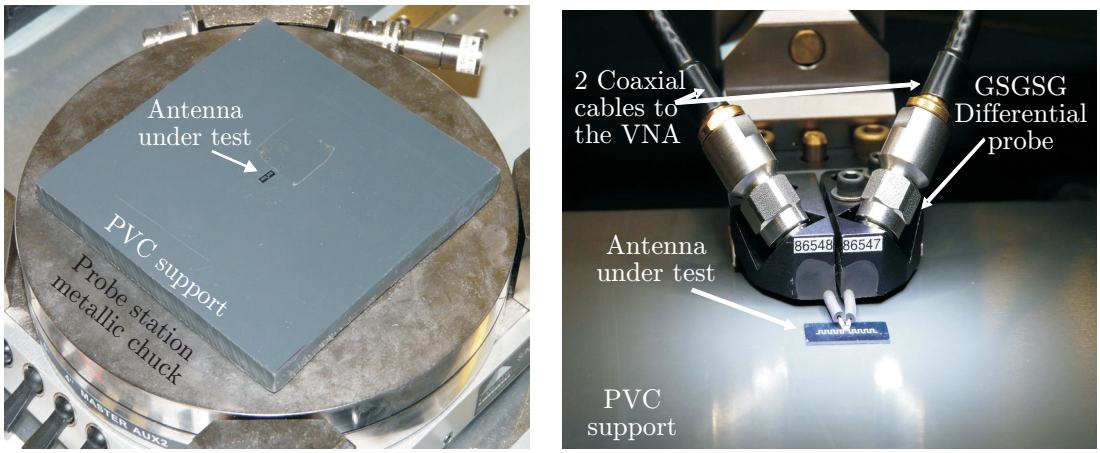


Figure 3.8: *Effects of the silicon conductivity in antenna parameters. Input reflection coefficient, input impedance and efficiency are shown. The antenna is a meandered dipole and chip dimensions are 5.25 mm x 1.9 mm.*

Ω·cm. Consequently, a conductivity value of 12.5 Ω·cm was used instead of the 20 Ω·cm of the wafer specifications. As can be seen in figure 3.8, the effects of conductivity are important: when conductivity is increased, radiation efficiency and gain are degraded and the impedance curves become smoother.

Dipole configurations with a differential feeding schema were used, taking 100 Ω as the differential reference impedance. A differential antenna topology was selected due to the fact that it can be easily connected to differential ended active circuits without the need for a balun. It is important to point out that good broadband matching levels were achieved due to the high silicon losses. Therefore, the main problem of integrated antennas is not the matching levels but the low gain and efficiency values.



(a) (b)

Figure 3.9: *Silicon antennas measurement setup. (a) A 1 cm PVC support was used to separate the antenna from the metallic chuck. (b) The differential dipole antennas were fed with a Picoprobe GSGSG probe.*

To validate the simulated results, measurements of the fabricated prototypes were carried out in a M150 Cascade Microtech probe station [96] using a Picoprobe GSGSG probe [97] to feed the differential dipole antennas. The antenna was separated from the probe station with a 1 cm PVC support. Otherwise the metallic chucks, directly in contact with the antenna, would invalidate the measured results. This PVC support was mechanized in order to transmit the vacuum to the antenna and avoid problems related to unwanted antenna movements during measurement. Figure 3.9 shows the measurement setup.

Dipole antennas with a differential feeding schema were measured. Typically, a balun is used to convert the single ended signal of the VNA into a differential signal. However, in the measurement of the frequency dependence of input parameters, the use of a balun can be a problem since broadband baluns are necessary. These baluns are expensive and can introduce errors. For this reason in the input parameters measurement we used a two port VNA and the mixed mode S parameters [98]. The differential input reflection coefficient can be calculated as

$$S_{dd} = \frac{(S_{11} - S_{21} - S_{12} + S_{22})}{2} \tag{3.1}$$

where S_{11}, S_{21}, S_{12} and S_{22} are the scattering parameters of the two single ended ports measurement. With this method, broadband measurements of differential antennas input parameters can be carried out avoiding the problems related to the baluns limited bandwidth

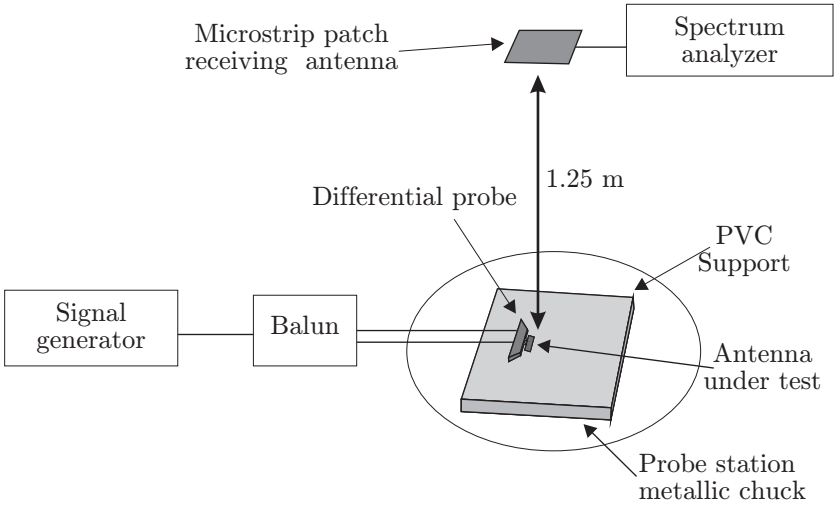
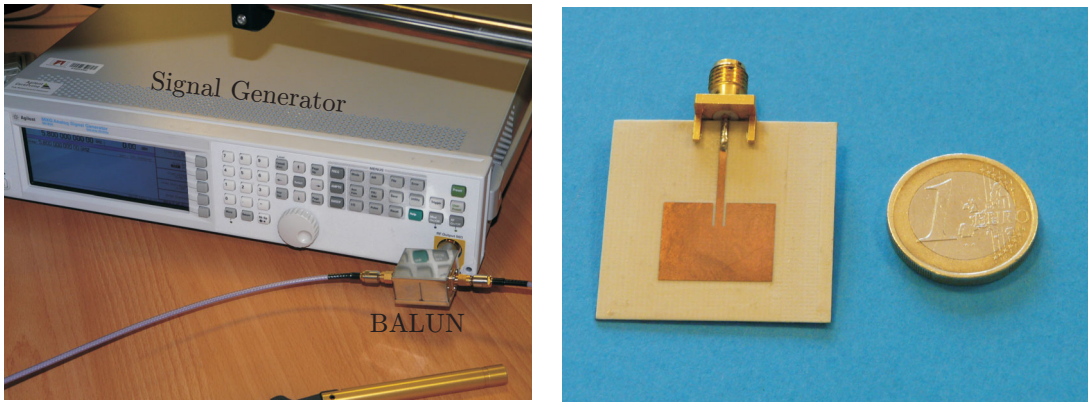


Figure 3.10: *Propagation measurement setup used for the gain measurement of the fabricated prototypes. Gain measurements were carried out in the 5.8 GHz ISM band.*

As previously said, one of the most limiting factors of the systems based on silicon integrated antennas are the low antenna gain values due to the high silicon losses. To obtain the fabricated prototypes gain, propagation measurements were carried out. In propagation measurements we could not use the mixed mode S parameters because a VNA with more than 2 ports was not available; therefore in these measurements we used a balun. Since these measurements were made at a single frequency we could use a simple microstrip narrowband balun.

Antenna gain was measured using the measurement setup shown in figure 3.10. A signal generator and a 5.8 GHz balun were used to feed the transmitting antenna (figure 3.11(a)). The receiving antenna was a 5.8 GHz microstrip patch built on Rogers RO4003C substrate ($\varepsilon_r = 3.38$, $\tan\delta = 0.0027$) (figure 3.11(b)). Measurements were performed inside the probe station, trying to maximize the clearance area near the integrated antenna and with a clear line of sight between the transmitting and the receiving antennas.

Although the probe station was adapted to minimize its effects on the antenna input parameters, it still affected the measurements. For this reason, two scenarios were considered in simulations. The first one was a simulation scenario that emulates the measurement setup: under the antenna there was a PVC block that represents the support and a PEC surface that models the effects of the probe station metallic chuck. In this scenario, good agreement between simulations and measurements was achieved (figure 3.12). Three identical antennas from different wafers were measured, proving the good repeatability of the fabrication process and measurements. In the second scenario the



(a) (b)

Figure 3.11: *Different components used for the propagation measurements. (a) Signal generator with a balun. (b) Receiving patch antenna fabricated on Rogers RO4003C substrate ($\varepsilon_r = 3.38$, $\tan\delta = 0.0027$).*

antenna was analyzed in free space in order to evaluate its performance when it is not affected by the probe station.

3.3 Miniaturization techniques applied to low resistivity silicon integrated antennas

The sensors will use the 5.8 GHz ISM frequency band to transmit the collected information to the central unit. At 5.8 GHz the wavelength is much larger than the chip dimensions, for this reason miniaturization techniques are necessary. In this section, the performance offered by various miniaturization techniques when they are applied to silicon integrated antennas will be analyzed. It is important to keep in mind that we will work with a high-loss substrate and this will affect the performance offered by these techniques. The achieved miniaturization factors will be presented, the effects on the radiation efficiency will be analyzed, and these results will be compared with the results obtained with a low-loss substrate.

3.3.1 Antenna designs

Different miniaturization techniques (figure 3.13) were used to obtain resonant dipoles around 5 - 6 GHz with a chip size of 12.5 mm x 2 mm. Two top-loading techniques were used, one based on metallic circles and the other based on capacitive loads. We also built several meandered dipoles with different number of meanders and strip widths (100 μm

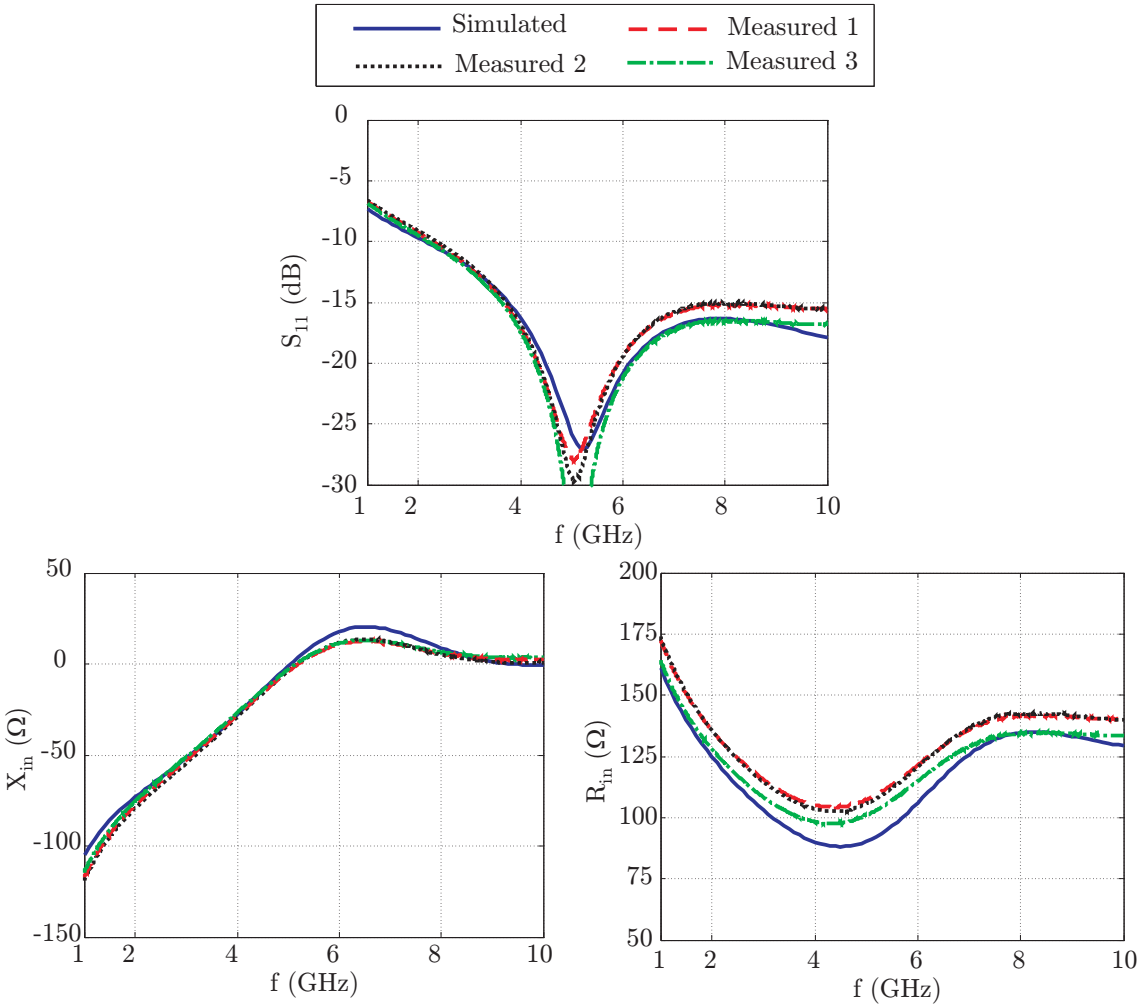


Figure 3.12: *Input parameters of a meandered dipole antenna fabricated on silicon. Solid line shows the simulated results. Dashed lines show the measured results of the same antenna fabricated on different wafers. A 100 Ω impedance was used as differential reference impedance.*

and 200 μm).

Figure 3.14 shows the simulated input parameters and the radiation efficiency of one of the meandered dipoles included in this study. The optimal matching point is near 5 GHz and the efficiency maximum is around 15 % at 5.3 GHz. This efficiency is higher than the efficiencies shown in figure 3.7 because the antenna of figure 3.7 was designed using the small chip size, and the antenna of figure 3.14 was designed using the large chip size. In the larger chips, less miniaturization is required, and as will be demonstrated later in this section, this is translated into an improvement in efficiency levels. A photograph of some of the fabricated prototypes can be seen in figure 3.15.

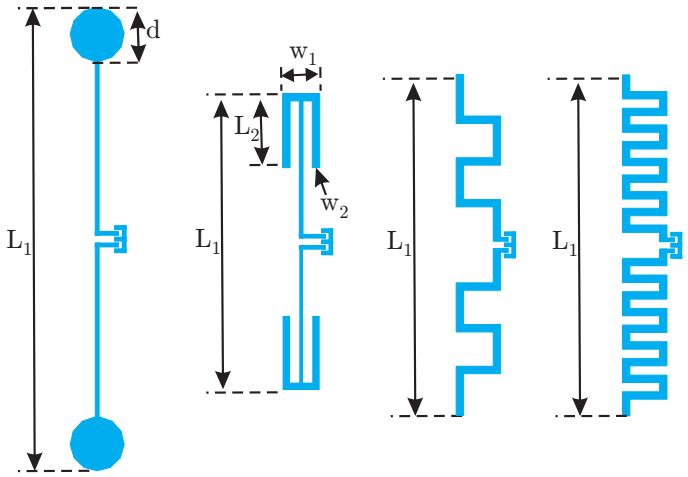


Figure 3.13: *Topologies included in the miniaturization study.*

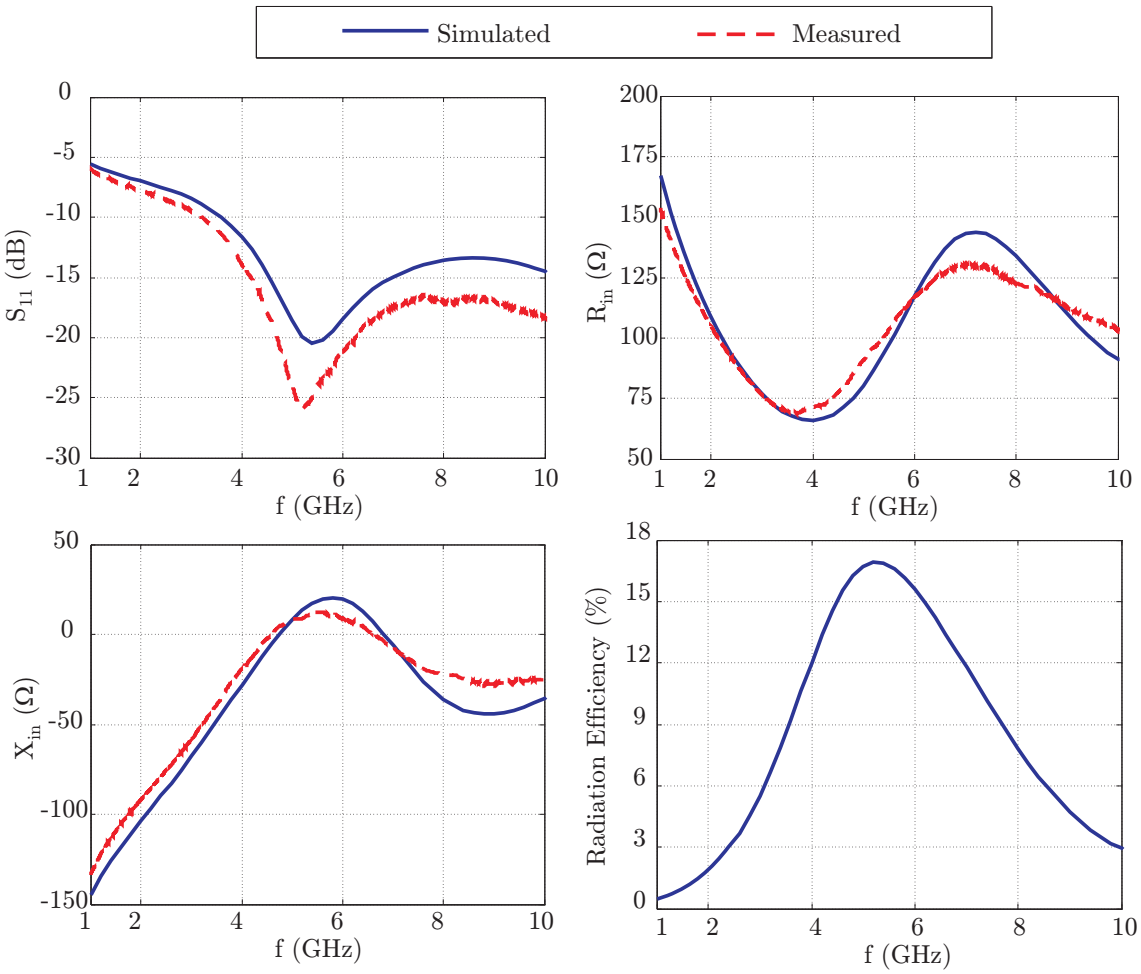


Figure 3.14: *Measured and simulated results of a meandered dipole (figure 3.13) with $N = 8$ meanders, conductor width = 100 μm and $L_1 = 9.2$ mm.*

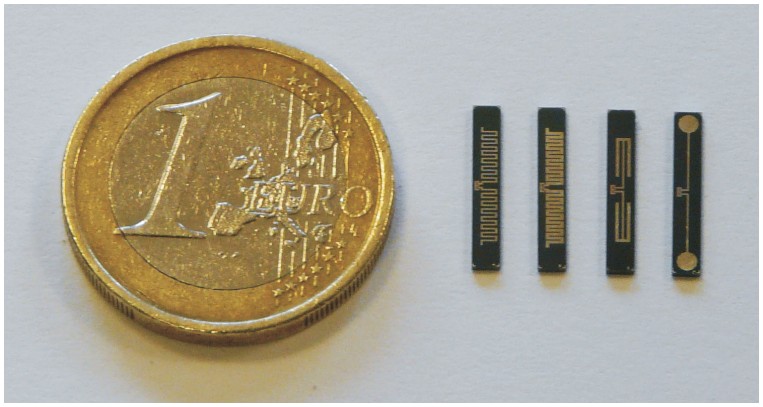


Figure 3.15: *Antennas with different miniaturization techniques. The chip dimensions are 12.5 mm x 2 mm.*

3.3.2 Miniaturization study: Measurements and simulation results

To compare the miniaturization offered by the different structures it is advisable to use the resonance frequency instead of the S_{11} minimum position since the former is not dependent on the reference impedance. Different dipole lengths (L_1) for the different topologies were used to guarantee resonance frequencies (f_{res}) of around 5 - 6 GHz, therefore they can not be compared directly. For this reason, the following figure of merit (F), valid for dipole configurations, is defined as

$$F = f_{res} \cdot L_1 \tag{3.2}$$

where it is assumed that the resonance frequency is inversely proportional to the dipole length. Low F values are desirable to achieve high miniaturization.

Using F, the miniaturization factor MF can be defined as

$$MF(\%) = 100\left(1 - \frac{F}{F_{flat_dipole}}\right) \tag{3.3}$$

where $F_{flat-dipole}$ is the measured figure of merit of a simple flat dipole built on silicon (reference antenna). The MF gives information about the length and area reductions (considering a constant chip width) offered by the different designs. High MF values are desirable.

The measured MFs obtained with the different topologies are plotted in figure 3.16(a). The reference antenna is a simple flat dipole with $F_{flat-dipole}$ = 68 GHz·mm. These miniaturization factors have been measured in the probe station scenario. Figure 3.16(b) shows the total simulated efficiency at the resonance frequency of the same topologies.

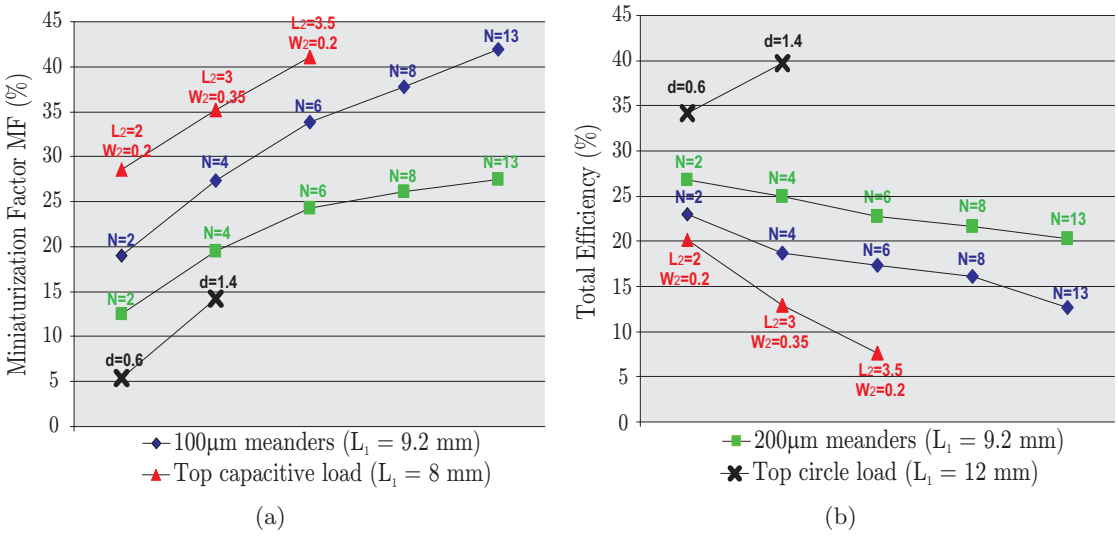


Figure 3.16: *Results of the miniaturization study in the probe station scenario. (a) Measured miniaturization factor (MF). (b) Simulated total efficiency of the different topologies. N is the number of meanders. The dimensions of the other variables (defined in figure 3.13) are in mm. The reference antenna for the MF calculation is a simple flat dipole with* $F_{flat-dipole} = 68$ *GHz·mm.*

In meandered dipoles resonance frequency decreases as the number of meanders increases. This is due to the increase in physical conductor length. However, when the number of meanders is high the efficiency of this miniaturization is reduced because close opposite currents tend to cancel each other out.

Measurements show that thin meandered dipoles (100 μm) provide more size reduction ($>$ 40 % of the total area), whereas miniaturization factors for thick meandered dipoles (200 μm) are smaller (27 %). As expected, when conductors are thin and the number of meanders (N) is increased, losses are higher and the efficiency is degraded. Therefore, there is a trade-off between antenna miniaturization and radiation efficiency which can be verified comparing figure 3.16(a) (positive slope as N increases) and figure 3.16(b) (negative slope as N increases).

Measured miniaturization factors for the dipoles with capacitive loads range from 30 % to 40 %, which are similar to the values obtained with the 100 μm meandered dipoles. Again the drawback is that radiation efficiency is reduced as the miniaturization factor is increased.

Circle loaded dipoles show the best efficiency levels, but the miniaturization obtained with these structures is less than 15 %. With this topology, there is not a clear trade-off between miniaturization and efficiency. When the loading circles diameter is increased,

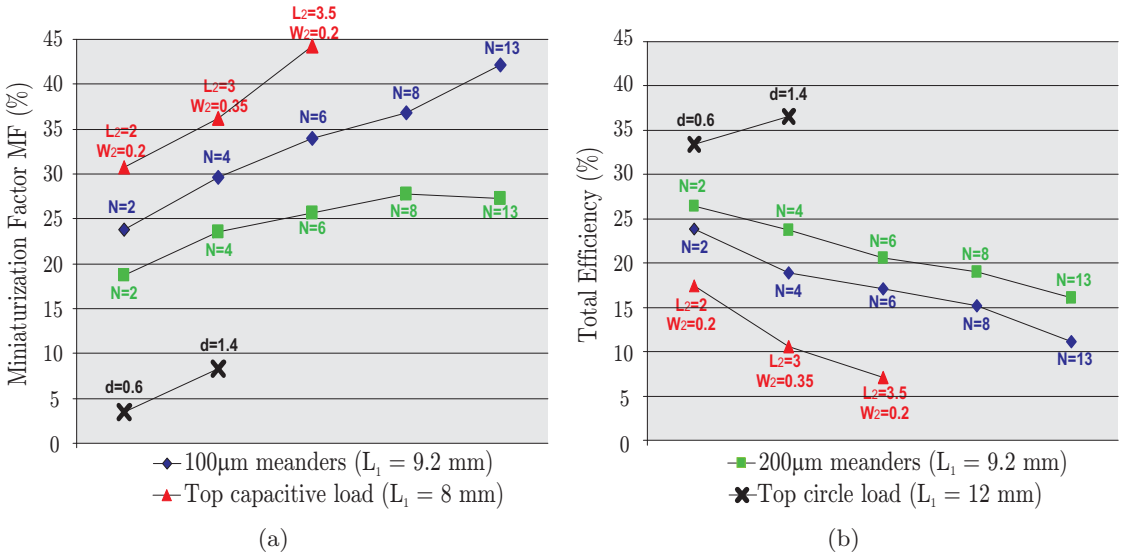


Figure 3.17: *Results of the miniaturization study in a free space scenario. (a) Simulated miniaturization factor (MF) (b) Simulated total efficiency of the different topologies. N is the number of meanders. The dimensions of the other variables (defined in figure 3.13) are in mm. The reference antenna for the MF calculation is a simple flat dipole with* $F_{flat-dipole} = 89$ *GHz·mm.*

both miniaturization and efficiency are increased, however it must be taken into account that the maximum diameter is limited by the chip width.

It is important to point out that the simulated results in a free space scenario (figure 3.17) are similar to those obtained in the probe station (checked by measurements). Hence, it is expected that similar results will be obtained in the real operating scenario of the antenna.

Radiation efficiency degradation when miniaturization techniques are applied is a well known effect. However, in silicon integrated antennas this efficiency degradation is much more significant than in low-loss substrate due to the high silicon conductivity. To prove this, the same antenna designs were simulated using a low-loss substrate with a similar dielectric constant (RO3010, $\varepsilon_r = 10.2$, $tan\delta = 0.0023$)[68].

The miniaturization factors obtained with this new substrate (figure 3.18(a)) are similar to the result obtained with the low resistivity silicon substrate (figure 3.16(a)). By comparing the radiation efficiency of the antennas built on the silicon substrate (figure 3.16(b)) and the efficiency of the antennas built on the low-loss substrate (figure 3.18(b)), it can be concluded that the efficiency degradation introduced by miniaturization techniques is emphasized when low resistivity substrates are used. Indeed, in the case of the thin meandered dipoles, in low-loss substrate the radiation efficiency is

degraded less than 15 % when the number of meanders is increased from 2 to 13, whereas in the case of standard CMOS silicon substrate the efficiency reduction is near 50 %. Table 3.1 shows the additional losses introduced by miniaturization. With the low-loss substrate a maximum degradation equal to 0.5 dB is observed, meanwhile when the silicon substrate is used, the degradation is higher than 2.5 dB.

Table 3.1: *Radiation efficiency and additional losses due to miniaturization of a 100 µm meandered dipole when the number of meanders is increased.*

		$N = 2$	$N = 4$	$N = 6$	$N = 8$	$N = 13$
Standard Silicon	Efficiency	23.0 %	18.6 %	17.4 %	16.0 %	12.7 %
	Additional losses with respect to $N = 2$	0 dB	0.8 dB	1.2 dB	1.5 dB	2.6 dB
Low-loss Substrate	Efficiency	93.3 %	90.8 %	88.7 %	86.2 %	82.7 %
	Additional losses with respect to $N = 2$	0 dB	0.1 dB	0.2 dB	0.3 dB	0.5 dB

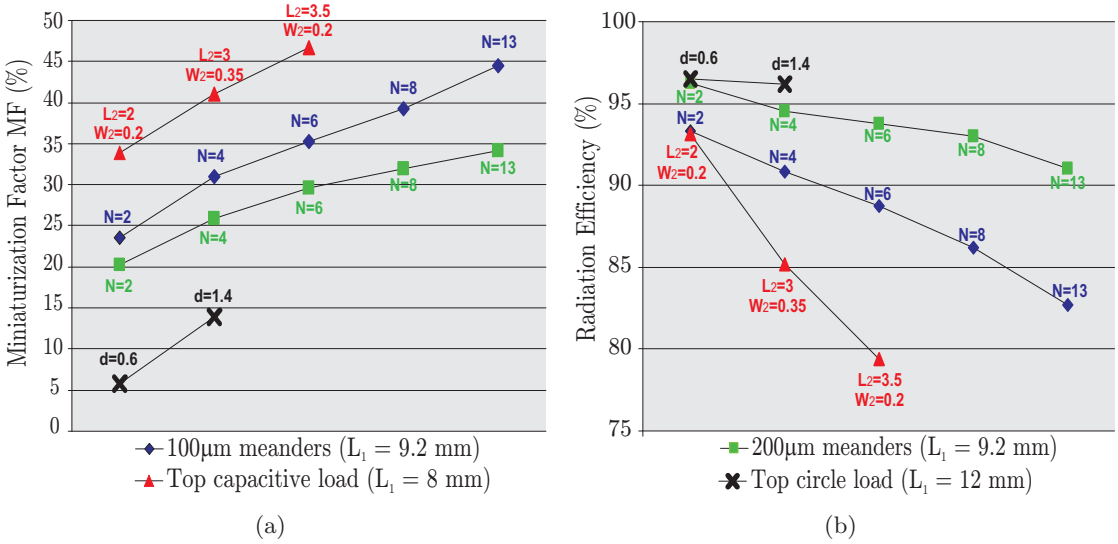


Figure 3.18: *Results of the miniaturization study in the probe station scenario when antennas are fabricated on a low-loss substrate (RO3010, $\varepsilon_r = 10.2$, $tan\delta = 0.0023$). (a) Simulated miniaturization factor (MF). (b) Simulated radiation efficiency of the different topologies. N is the number of meanders. The dimensions of the other variables (defined in figure 3.13) are in mm. The reference antenna for the MF calculation is a simple flat dipole with $F_{flat-dipole} = 75.5$ GHz·mm.*

3.4 Improving silicon integrated antennas by substrate micromachining

As it has been shown in the previous section, one of the main drawbacks of antenna integration are the low radiation efficiency levels due to the silicon losses and the efficiency degradation introduced by miniaturization techniques. In literature several solutions to improve antenna efficiency can be found, such as the use of a higher substrate resistivity, addition of low-loss insulating layers, substrate thinning or selective removal of the substrate by means of micromachining.

For instance, in [99] proton implantation of selected substrate areas is used to increase the silicon resistivity. The main problem of this technique is that rare special equipments are necessary. Another solution is to increase the distance between the radiating element and the lossy silicon. Thicker SiO_2 layers [81] or an additional layer of a low-loss material such as benzocyclobutene (BCB) can be used to insulate the antenna. These materials have low dielectric constants and very low losses, but are difficult to apply with a thickness greater than 20 - 30 μm, therefore efficiency improvements are limited. In [80, 87, 100] different examples of application of these materials as insulation layers can be found.

In [84] the silicon substrate thickness is reduced to improve the on-chip antenna efficiency. Another solution is to apply micromachining techniques to selectively etch silicon of the back side of the antenna. For example, [94] uses a micromachining process to improve the efficiency of a dipole antenna in the 5.8 GHz band. Other examples of micromachined antennas can be found in [87, 88, 101]

In previously reported designs [87, 94] a single rectangular shaped mask has been used for the selective silicon etching process. In this section, we will carry out a comparative study of several etching patterns in order to maximize the gain improvement without a significant increase in the antenna fragility. The antenna performance will be analyzed in various operating scenarios to prove that similar gain improvements can be expected in real applications.

3.4.1 Antenna designs, definition of the etching areas and fabrication process

The radiating element is a meandered dipole designed to work at frequencies near 5 GHz (figure 3.19(a)). The small chip dimensions (1.9 mm x 5.25 mm) were used for the prototypes fabrication.

The best option to improve the radiation efficiency would be to remove all the silicon

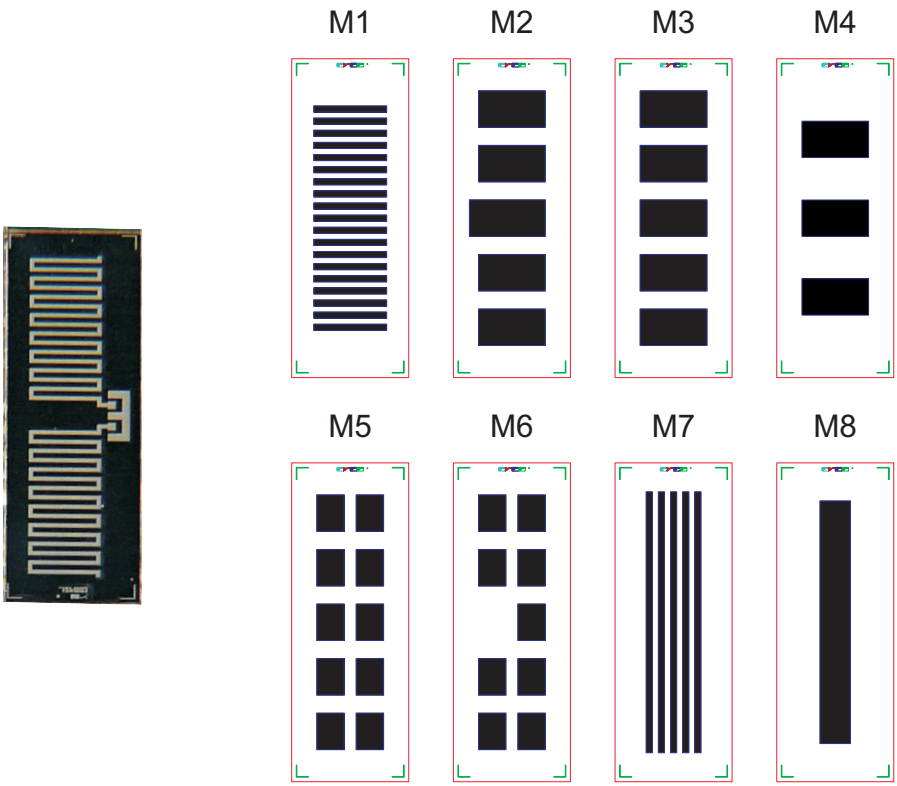
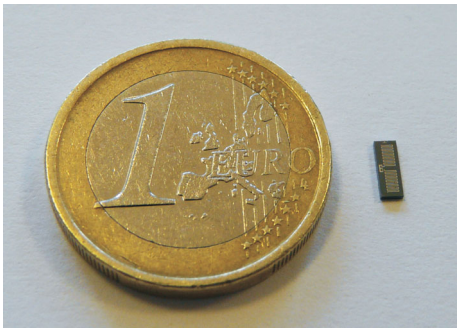


Figure 3.19: *Design of the fabricated prototype antennas. (a) Image of the fabricated meandered dipole antenna. (b) Schematic representation of the etching patterns used for the micromachining process. Black rectangles indicate the areas where silicon is etched.*

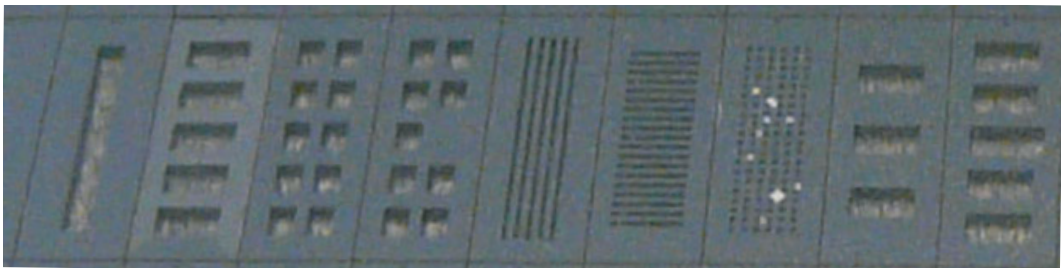
under the antenna, but the etching of large areas is problematic because the resulting membrane structure is fragile and can be easily broken. Etching an array of smaller areas can provide similar improvements without compromising the mechanical robustness. The etching patterns under study (figure 3.19(b)) try to avoid the elimination of a large, continuous silicon area, keeping silicon walls to guarantee the structure robustness.

Fabrication started with p-type silicon wafers having a resistivity of 20 Ohm·cm and a thickness of 500 μm. A 2 μm-thick layer of SiO_2 was grown with a wet oxidation process and a 4 μm-thick SiO_2 was deposited by Plasma Enhanced Chemical Vapor Deposition (PECVD). The resulting thickness of SiO_2 (6 μm) is similar to the effective thickness of the dielectric stack that some CMOS technologies have under the top metal level, so the results of the present study are valid for antennas made with standard technologies.

Next, a 2 μm-thick aluminum (Al) was deposited by sputtering. The Al layer was patterned by standard photolithography and wet etch. On the back side of the wafers, a



(a)



(b)

Figure 3.20: *Prototype antennas with micromachining. (a) Top view of one of the antennas. (b) Rear view of some of the antennas.*

1 μm-thick Al layer was deposited and patterned in the same way as above. Subsequently, the SiO_2 from the back side was removed with a dry etching process and a Deep-Reactive Ion Etching (DRIE) process was used to partially remove the silicon beneath the antenna structures. The fabrication was finished by removing the back side Al with a wet etching process. More information about these fabrication processes can be found in [102]. Figure 3.20 shows some of the fabricated prototypes.

3.4.2 Performance improvement due to the micromachining process: Measurements and simulation results

In figure 3.21 the input reflection coefficient of the antenna with different etching patterns is plotted. A slight frequency shift to higher frequencies can be seen due to the fact that the elimination of silicon reduces the effective relative permittivity. When micromachining is applied impedance curves also become slightly sharper due to the lower losses, reducing the matching bandwidth of the antenna. The effects of the different micromachining configurations on the input parameters are similar. From figure 3.21(b) it is seen that the above mentioned effects become more important when the etching depth is increased.

Micromachining effects on the input reflection coefficient are not so important to

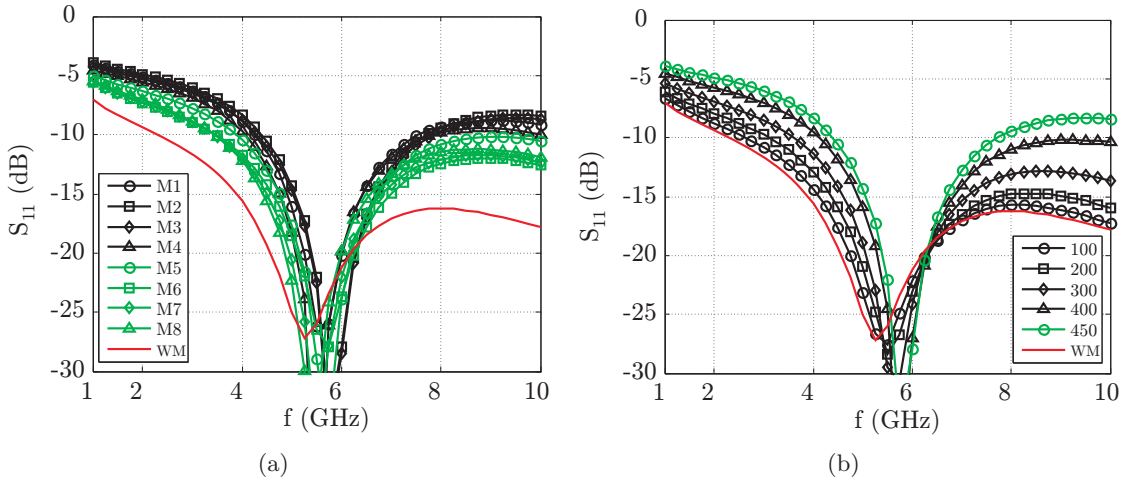


Figure 3.21: *Simulated input reflection coefficient. (a) Different etching configurations are applied. The etching depth is 450 μm. (b) Different etching depths (μm) are applied. The M2 pattern is used for the micromachining process. The wafers thickness is 500 μm. WM = Without Micromachining.*

motivate the use of these techniques; it is in the antenna radiation parameters where a significant performance improvement is obtained. In figure 3.22(a) the effects of the etching process in the radiation efficiency can be clearly seen because an efficiency improvement is observed when micromachining is applied. Depending on the etching pattern, this improvement can be higher than a factor of 2. It is important to point out that our band of interest (5.8 GHz) is near the efficiency maximum.

Figure 3.22(b) shows the efficiency at 5.8 GHz for the different etching patterns as a function of the etching depth. The wafers thickness is 500 μm in all cases. The efficiency improves when the depth is increased, and again, this improvement also depends on the geometry of the etched areas. Similar conclusions can be obtained from figure 3.22(c). A gain improvement of up to 4 dB can be achieved in the normal direction with the best etching pattern (M1). For the same etching depth (i.e. 450 μm), the gain improvement can range from 1.5 dB to 4 dB depending on the etching pattern used in the micromachining process.

To verify these results some gain measurements of the fabricated prototypes were carried out. The measured and simulated gain improvements are summarized in table 3.2. Although measured gains are slightly higher than simulator predictions there is a good agreement in the trends. With an etching depth of 450 μm measured gain improvements of near 4 dB are obtained. Higher etching depths would improve the gain even more, but the resulting structures would be more fragile. With 50 μm of silicon in the cavities, the resulting structures are strong enough to withstand the measurements in the probe station

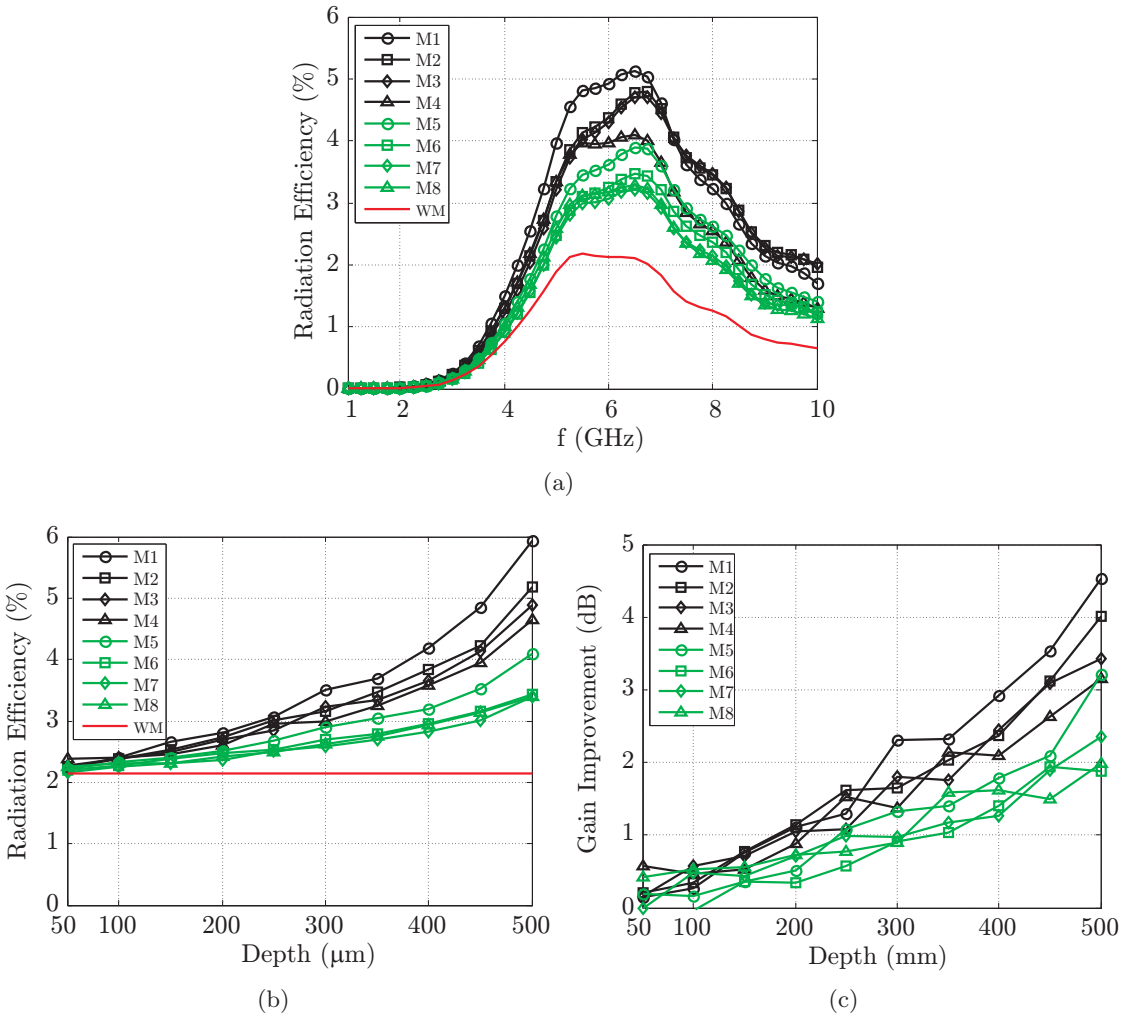


Figure 3.22: *Simulated radiation parameters when micromachining techniques are applied. (a) Radiation efficiency as a function of frequency. The etching depth is 450 μm. (b) and (c) radiation efficiency and gain improvement as a function of the etching depth. The operating frequency is 5.8 GHz and the wafers thickness is 500 μm. The reference gain for the gain improvement plot is the gain of the antenna without micromachining (−18.95 dB). WM = Without Micromachining.*

Table 3.2: *Measured and simulated gain improvement for the different etching patterns. The last column shows the etched area. The etching depth is 450 μm and the wafers thickness is 500 μm.*

Etching pattern	Simulated Gain Improvement	Measured Gain Improvement	Etched Area (mm^2)
M1	3.5	3.9	2.28
M2	3.1	3.6	3.39
M3	3.1	3.3	3.30
M4	2.6	2.8	1.98
M5	2.1	2.6	2.70
M6	1.9	2.4	2.43
M7	1.9	2.0	2.15
M8	1.5	1.8	2.00

without the need to fill the cavities with another dielectric material. We tried to completely remove the silicon from the patterns but in some of the prototypes the remaining dielectric membrane supporting the antenna broke, probably due to internal mechanical stress in the oxide layers.

Intuitively, it could be deduced that the best etching pattern is the one that eliminates more silicon under the antenna. However, this is only partially true because the geometry used for the etching process is also important. According to table 3.2, the geometry with the highest gain improvement is not the one that eliminates most silicon. The amount of silicon etched in M2 is 50 % higher than in M1 but the measured gain improvement in M1 is 0.3 dB better than in M2. In a similar way, the amount of silicon etched in M4 and M8 is nearly the same, but M4 gain improvement is 1 dB better than M8.

Based on these results, some general guidelines for the use of micromachining techniques can be given in order to maximize the gain improvement of silicon integrated dipole antennas. First of all, structures with horizontal (normal to the dipole orientation) trenches show better performance than structures with vertical ones. This is a reasonable result because horizontal trenches limit the current circulation in the substrate.

Secondly, it is better to eliminate silicon near the feeding area, where currents and fields are stronger than in the areas far away from the feeding point. This can be verified in figure 3.23, where only a small part of the silicon chip around the position Δy is etched. When the etched area is under the feeding, gain is up to 2 dB better than when it is under the antenna ends.

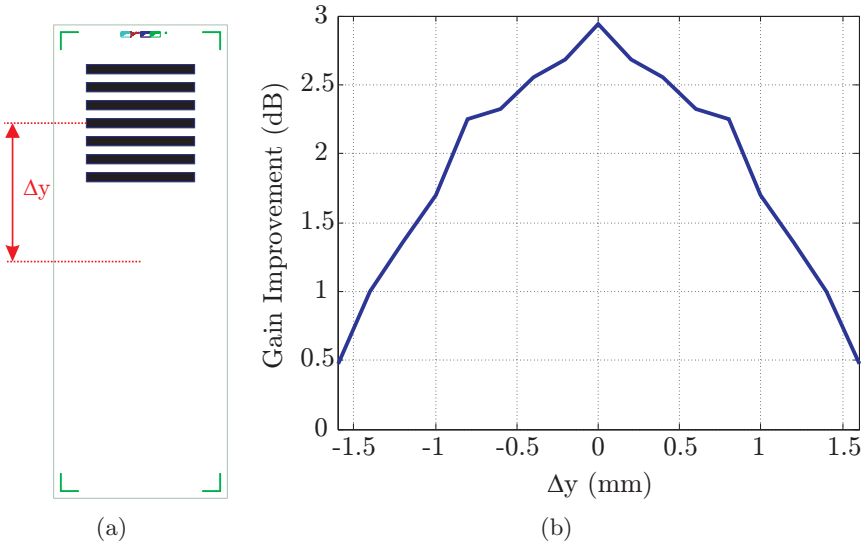


Figure 3.23: *Simulated gain improvement as a function of the etched area position. (a) Pattern used in the micromachining process. (b) Gain improvement plot. The reference gain is the gain of the antenna without micromachining (−18.95 dB).*

3.4.3 Antenna performance improvement due to micromachining in a free space scenario

In the previous section the micromachining effects have been analyzed through simulations and measurements carried out inside the probe station. One could think that the obtained results only apply in that particular case. In this section, simulations are used to assess the antenna performance in free space.

Comparing the S_{11} parameters in a free space scenario (figure 3.24(a)) with the results of the previous section (figure 3.21), the main difference is a slight frequency shift to higher frequencies. This is related to the fact that, in the probe station scenario, the antenna is on a supporting material with a relative permittivity higher than one, which tends to reduce the resonance frequency.

The gain improvement at 5.8 GHz with respect to an antenna without micromachining is shown in figure 3.24(b). The improvement is similar to the values obtained in the probe station scenario. Comparing figure 3.22 and figure 3.24, it can be seen that if the etching patterns are sorted according to their gain improvement, the same result is obtained in both scenarios. Consequently, it is expected that this improvement will also be valid in the real operating scenario of the antenna which will depend on the final application.

In this section it has been shown that with micromachining techniques a gain improvement near 4 dB is obtained. This gain improvement is relevant. If these antennas

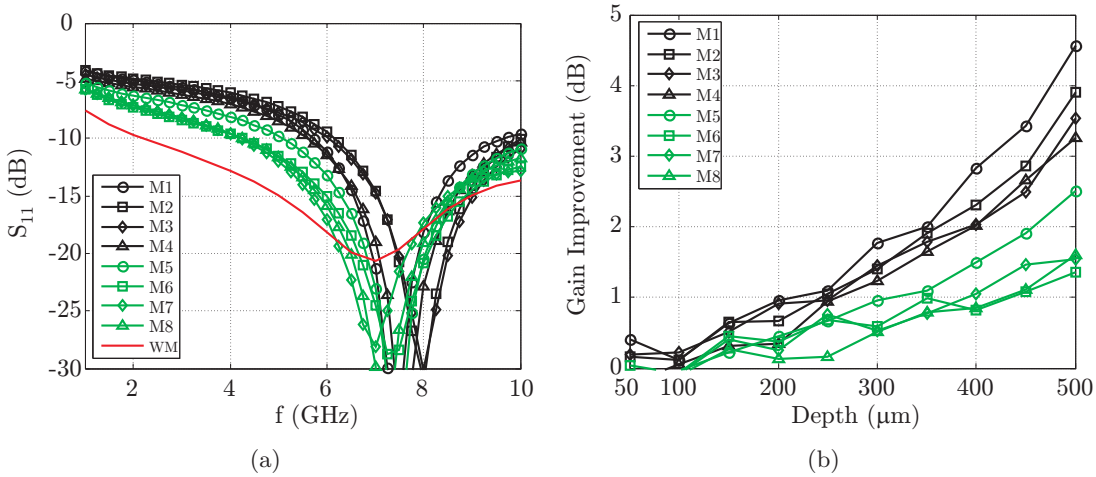


Figure 3.24: *Performance improvement due to micromachining in a free space scenario. (a) Simulated input reflection coefficient. The etching depth is 450 µm. (b) Gain improvement as a function of the etching depth. The operating frequency is 5.8 GHz and the wafers thickness is 500 µm. WM = Without Micromachining.*

are used in our wireless sensor network, the power required for the communication between nodes (keeping the same system performance) can be reduced by more than fifty percent thus increasing battery life. These techniques can be also useful to mitigate the additional losses introduced when miniaturization techniques are applied to reduce antenna dimensions.

3.5 Antenna performance degradation due to the presence of coils for an inductive link energy coupling

According to [103], previously presented remote and implantable sensors are relatively large due to the use of a sizeable battery for powering the sensor for a long time. To solve this problem it would be of interest to have a way to transmit energy to the sensor and sporadically reload its battery without having physical access to it, thus allowing for smaller batteries. In the normal system operation mode the remote sensor continuously sends information to an external receiver using the RF link (a range of over a meter [104]), but when the battery is flat, it can be reloaded with a short range inductive coupling (a range of a few centimeters [103]).

The main drawback of this configuration is that the presence of the coil for the low frequency inductive coupling will significantly degrade the RF integrated antenna performance. The coil is usually one of the bigger chip elements, and hence one of the elements that can introduce higher degradation levels to the RF link. This section will be

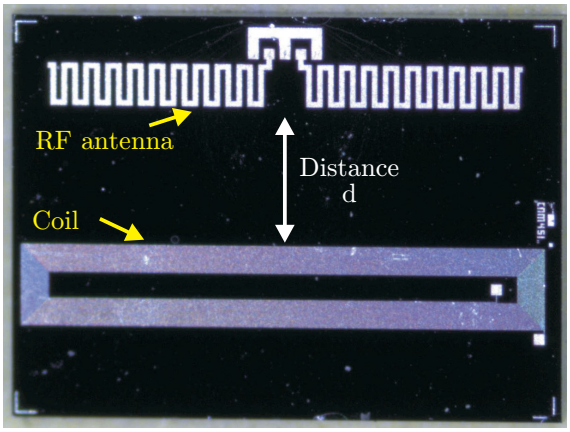


Figure 3.25: *Fabricated prototype antenna for the evaluation of the coil effects on the RF link. The chip dimensions are 3.8 mm x 5.2 mm.*

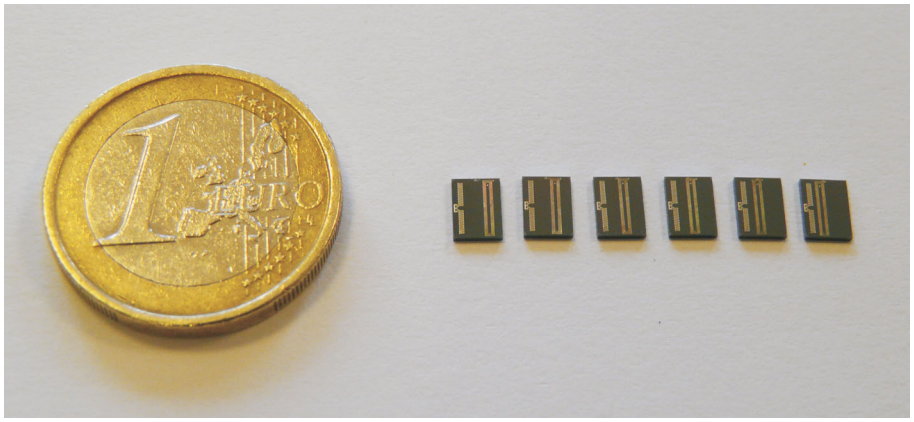


Figure 3.26: *Prototype antennas. The chip dimensions are 3.8 mm x 5.2 mm. The distances d between the RF antenna and the coil are 320 µm, 570 µm, 820 µm, 1.07 mm, 1.32 mm and 1.57 mm.*

focused on the study of the effects of this coil. One of the main parameters that defines the performance degradation in the RF link is the distance d between the coil and the RF antenna (figure 3.25). For this reason several prototypes changing this distance d were fabricated and measured (figure 3.26). As can be seen in these figures, the radiating element is a meandered dipole. The chip dimensions are 3.8 mm x 5.2 mm.

3.5.1 Performance degradation due to the coil effects: measurements and simulation results

In spite of the presence of the coil, antennas are well matched to a 100 Ω differential reference impedance (figure 3.27), showing that the coil does not significantly change the antenna input parameters. Where the degradation is more significant, is in the radiation

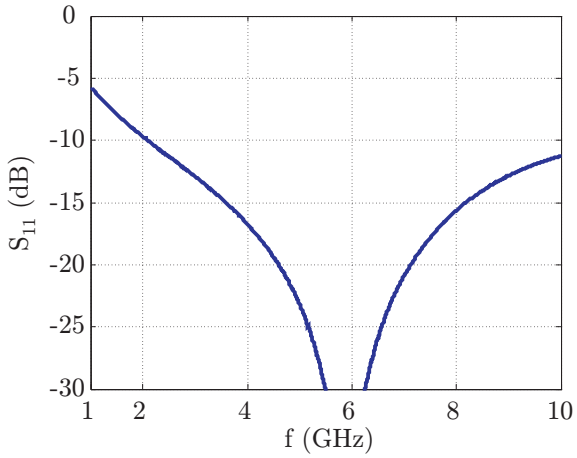


Figure 3.27: *Measured input reflection coefficient of an integrated antenna fabricated together with a coil for an inductive link (figure 3.25). The distance between the antenna and the coil is $d = 320\ \mu m$.*

parameters.

Figure 3.28(a) shows the simulated and measured gain of the fabricated prototypes. The x-axis represents the separation d between the antenna and the coil, ranging from 320 μm to 1.57 mm. Good agreement between simulations and measurements is achieved. Measurements were repeated twice, showing that the measurement setup offers good repeatability in the results. The measured gain when the coil is at the maximum distance

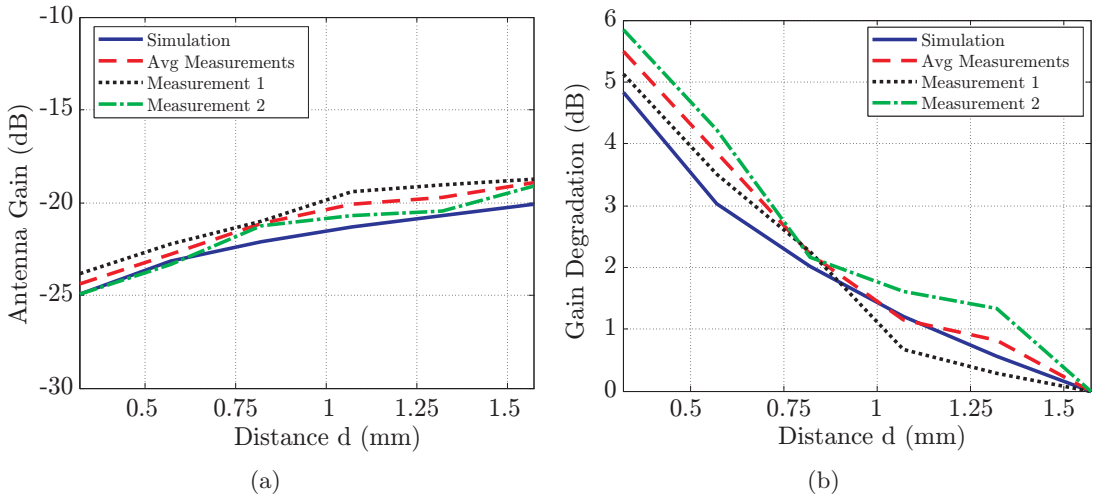


Figure 3.28: *Radiation parameters as a function of the distance d between the RF antenna and the coil. (a) Antenna gain. The simulated gain without the coil is -17.7 dB. (b) Gain degradation. The reference gain is the gain when the coil is at $d = 1.57$ mm. Operating frequency $f = 5.8$ GHz.*

Table 3.3: *Measured gain degradation and system range reduction as a function of the distance d between the RF antenna and the coil.*

Distance d (mm)	Gain degradation (dB)	System range reduction (%)
0.32	5.5	47
0.57	3.9	36
0.82	2.2	23
1.07	1.2	12
1.32	0.8	9
1.57	–	–

($d = 1.57$ mm) is near -19 dB and the gain when the coil is closest to the antenna is near -24.5 dB. The simulation of the antenna without the coil gives a gain value near -17.7 dB.

To reduce the total chip dimensions, it is desirable to have the coil close to the antenna, however when the distance between both elements is reduced the gain degradation is more significant (figure 3.28(b)). Table 3.3 summarizes the measured performance degradation when the coil approaches the antenna. Taking a distance of 1.57 mm as reference, a degradation of 2.2 dB can be seen as this distance is reduced to 820 μm. If the distance is further reduced to obtain a smaller chip, the gain degradation can be higher, obtaining 5.5 dB when the distance is equal to 320 μm.

In the last column of table 3.3 the system range reduction has been calculated assuming a free space propagation model. All the other system parameters are kept constant when the separation d is changed. Reductions of nearly 50 % of the system range can be observed. These results prove that when the inductive coil and the RF antenna are integrated in the same chip, there will be a clear trade-off between chip dimensions (limited by the separation between the coil and the RF antenna) and the system performance.

3.5.2 Performance degradation due to the coil effects in a free space scenario

In this section, the coil effects when the antenna is in a free space scenario will be analyzed by means of simulations.

In this scenario, the simulated antenna gain without the coil is equal to -21.6 dB. In the probe station this gain is higher because the support and the metallic chuck modify the radiation pattern shape increasing the gain of the dipole in the broadside direction. Figure 3.29 shows the simulated gain degradation due to the coil effects in a free space scenario as a function of the distance d between the coil and the RF antenna. Despite

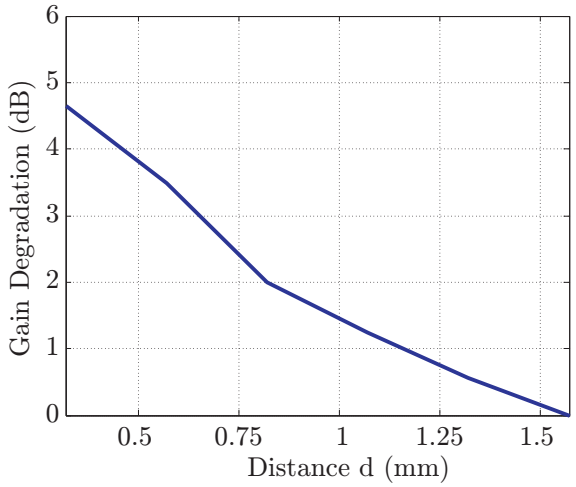


Figure 3.29: *Simulated gain degradation in a free space scenario as a function of the distance d between the RF antenna and the coil. The reference gain (-25.5 dB) is the gain when the coil is at d = 1.57 mm. Operating frequency f = 5.8 GHz.*

the fact that the absolute gain values are not the same in both scenarios, the degradation levels due to the presence of the coil are similar. Therefore it is predictable that similar degradation levels will be also obtained in the operating scenario of the real application.

3.5.3 Effects of the performance degradation in a real system link budget

Now, we will show the coil effects on the performance of a real communication system: a remote sensor network similar to the system that has inspired this thesis. The system parameters will be based on the results presented by Lin in [75]. However, we are working with low power sensors and for this reason the transmitted power will be lower than the value used in [75] in order to reduce the power consumption. The receiver antenna will be a patch antenna with a gain of 6 dB. System parameters are summarized in table 3.4.

With the information of table 3.4, and assuming a free space propagation model, the

Table 3.4: *System parameters used for the link budget.*

Link Margin	12 dB
Tx Power	-10 dBm
Rx Antenna Gain	6 dB
Rx Sensitivity	-97.8 dBm
Thermal Noise	-173.8 dBm/Hz
Data rate (100 kbits/sec)	50 dBHz
Eb/No	18 dB
Rx Noise Figure	8 dB

Table 3.5: *System range for different chip configurations.*

Antenna	Antenna measured gain (dB)	System Range (m)
Without coil	-17.7	6.6
With the coil at 1.57 mm	-19.0	5.7
With the coil at 320 μm	-24.5	3

values of table 3.5 have been obtained. The original RF link range (without the parasitic coil) is greater than 6.5 m. However when the coil is integrated with the antenna in the same chip, this range is reduced to 5.7 m, and if the distance between the antenna and the coil is reduced to 320 μm, the system range is further degraded to 3 m.

In this section we have studied the effects of a coil, but the same conclusions are also valid for any other element integrated with the antenna, such as the sensor or the other electronic circuits, because in the real application the antenna will not be alone in the chip. Furthermore it is important to point out that, as was shown in the previous section, this system performance degradation could be mitigated by the use of micromachining techniques to improve the antenna radiation efficiency.

3.6 Chapter conclusions

In this chapter silicon integrated antennas have been presented as a possible solution for the remote sensor antenna. Silicon integrated antennas allow the integration of all the elements into a single chip, obtaining a real system-on-chip solution and reducing the packaging costs and system dimensions. To minimize costs and make the integration feasible, the antenna fabrication process must be compatible with the standard CMOS technologies. This means that low resistivity silicon must be used as substrate, which significantly degrades the antenna radiation efficiency.

The communication between the sensors and the central unit uses the 5.8 GHz ISM band. At this frequency antennas are big and difficult to integrate, and hence miniaturization techniques must be applied. With some simple techniques, like meandered lines or loaded structures, size reductions near 40 % can be obtained. However, it must be taken into account that with the low resistivity silicon substrates used for the antennas fabrication, the trade-off between miniaturization and efficiency degradation becomes much more significant.

Low efficiency levels could be defined as the main problem of silicon integrated antennas. The use of micromachining techniques to eliminate part of the silicon under the

antenna is a good solution to improve these efficiencies. The micromachining effects are more significant in the radiation parameters than in the input parameters, and antenna gain improvements near 4 dB have been measured. The higher the etching depth, the better the efficiency improvement. However, when the depth is increased the structure is more fragile and the fabrication process is more difficult (we had fabrication problems with etching depths greater than 450 μm).

In the micromachining process, not only the total amount of etched silicon is important, but the way this silicon is eliminated is also significant. The most sensitive area is the area near the feeding point, where generally fields and currents are stronger, for this reason it is important to increase the amount of etched silicon in this area. Furthermore, measured and simulated results show that the configurations with trenches perpendicular to the dipole orientation offer better gain improvement. Measurements show that with the same etching depth (450 μm), gain differences greater than 2 dB can be observed depending on the etching pattern.

The effects of this gain improvement can be relevant to system performance. For instance if these antennas are used in the sensors of the sensor network, keeping the same system performance, the power delivered by the sensor to the antenna can be reduced by more than fifty percent when micromachining techniques are applied, significantly increasing battery life.

In real applications, the antenna will be integrated together with other elements built into the same chip. Normally these elements are small, but in some cases, their dimensions can be significant, and therefore, the presence of these "parasitic" elements will degrade the RF antenna performance. In this chapter, the effects of having a coil for an inductive energy coupling near the antenna have been analyzed. The presence of the coil degrades the antenna gain taking into account that when the separation between both elements is reduced the chip can be smaller, but the performance degradation is higher. Reductions of 50 % in the system range have been measured as this separation is reduced from 1.57 mm to 320 μm.

To sum up, the results of this chapter demonstrate that silicon integrated antennas in the 5.8 GHz ISM frequency band are a feasible solution and that they can be fabricated using standard CMOS technologies together with other elements such as a coil or a sensor. Therefore, silicon integrated antennas are suitable for the sensors of our sensor network. The main problem is the low efficiency levels that degrade the system performance, however measurements show that these efficiency levels can be improved with micromachining techniques.

Chapter 4

Human body effects on antennas performance

In a remote sensor network, some of the wireless sensors could be implanted inside the body. The human body is a complex dielectric structure of materials with high conductivity and permittivity values that can adversely affect antenna performance. So the study of the body tissues effects is extremely important to ensure the correct operation of wireless communication systems operating in the vicinity of or within the body.

This chapter will present a review of previous works and existing regulations. Then, the problems related to simulation and measurement of the body effects on antennas performance will be discussed and a detailed study of the propagation losses introduced by the body tissues in different ISM bands will be presented. The differences between several body models will be also analyzed. Finally, measurements with body tissue simulating liquids will be carried out to demonstrate the simulations validity.

4.1 Introduction to the human body effects

There exists a great diversity of applications where antennas interact with human tissues, but basically two scenarios can be defined: The case where both antennas are outside the body, and the case where one of the antennas is implanted inside the body. This section will present a review of previous works and existing regulations for both scenarios.

4.1.1 Antennas in the vicinity of the human body

During the last decade there has been an important increase in public awareness of

radiation hazards associated with mobile phones and in particular, with base stations. Consequently, multiple references analyzing the possible effects of cellular base stations on humans have appeared. As an example, recently the Spanish Ministry of Industry, Tourism and Commerce, and the College of Telecommunication Engineers (COIT) have jointly published a leaflet [105] providing some safety guidelines related to cellular base stations, in an attempt to answer the questions that people typically ask about the possible effects of base station radiation on the human body.

[106] investigates the minimum necessary distance between a human body and a cellular base station in order to comply with the threshold values of the Institute of Electrical and Electronic Engineers (IEEE) and the International Commission on Non-Ionizing Radiation Protection (ICNIRP). In [107] human exposure to electromagnetic fields is analyzed in a realistic urban environment, incorporating possible signal reflections of walls and buildings.

The analysis of the mobile terminal effects on the head tissues is also an important issue because the separation between the mobile phone and the users head during a call is minimal. [108] shows an extensive study of the electromagnetic absorption into the human head and neck due to mobile phones radiating at 835 MHz and 1900 MHz. The fields for various head models (adult and child head) are analyzed and compared with the regulation limits. Other references where the mobile phones effects are studied can be found in [109–113], where different antenna configurations are considered.

Another field of interest where antennas are close to the human body are wearable devices [114] and Body Area Networks (BAN). In these cases the propagation channel is quite different from a free space propagation model. For this reason, there are multiple references where the problem of BAN channel modeling is analyzed [115–117]. Another sensitive issue is the design of the antenna because small, flexible and low profile antennas are usually required to improve user mobility and comfort. Textile antennas made of electro-textile materials are a good solution for these applications [118–120].

In the previous examples the antenna is used as a radiating element for information transmission. However, in some applications, like breast tumor detection, antennas near the body are used with imaging purposes [121–124].

Guidelines and regulations for antennas in the vicinity of the human body

In certain European countries a social concern has been raised about the possible health hazards resulting from RF fields exposure. As a consequence, several international radiation safety standards, such as the European Commission Recommendation related to electromagnetic exposure (ERC, July 1999 [125]), have been published. Most of them are

based on the ICNIRP guidelines [126] and the IEEE standard of 1999 [127] (updated in [128] and [129]). In the United States, the Federal Communication Commission (FCC) has provided some guidelines to assist local government officials in devising efficient procedures to ensure that antennas comply with FCC limits for human exposure to radiofrequency [10].

In the frequency range from 100 KHz to 10 GHz, restrictions are generally based on the Specific energy Absorption Rate (SAR) values in order to prevent excessive and harmful heating of body tissues [125]. The European Commission gives the following definition: "The Specific energy Absorption Rate (SAR) averaged over the whole body or over parts of the body is defined as the rate at which energy is absorbed per mass unit of body tissue and it is expressed in watts per kilogram (W/kg). Whole body SAR is a widely accepted measure for relating adverse thermal effects to RF exposure. Besides the whole body average SAR, local SAR values are necessary to evaluate and limit excessive energy deposition in small parts of the body resulting from special exposure conditions. All SAR values are to be averaged over any six minute period" [125].

This definition can be expressed as [130]

$$SAR = \frac{d}{dt}\left(\frac{dW}{dm}\right) = \frac{d}{dt}\left(\frac{dW}{\rho dV}\right) \tag{4.1}$$

where dW is the incremental energy absorbed by (dissipated in) an incremental mass (dm) contained in a volume element (dV) of a given density (ρ) .

The SAR values are related to the electric field at a given point by [130]

$$SAR = \frac{\sigma |E|^2}{\rho} \tag{4.2}$$

where σ is the tissue conductivity and E is the RMS electric field strength. Finally, under ideal non-thermodynamic circumstances, SAR can be related to the increase in temperature by [130]

$$SAR = \frac{c \triangle T}{\triangle t} \tag{4.3}$$

where $\triangle T$ is the change in temperature, $\triangle t$ is the duration of exposure and c is the specific heat capacity.

The SAR limitations according to the ICNIRP and IEEE guidelines are summarized in table 4.1. These limitations are mainly based on the body temperature rise due to exposure, taking into account that a maximum increase lower than 1^oC is desirable. The IEEE standard of 1999 [127] presented some local SAR limits different from the ICNIRP guidelines, but in the 2005 revision of the standard, the peak spatial average SAR values

Table 4.1: *SAR: Whole body and maximum local limits [125–128].*

Exposure Situation	Whole Body Average SAR (W/Kg)	Local SAR (Head and trunk) (W/kg)
Occupational	0.4	10
General Public	0.08	2

were changed with the aim of harmonizing the basic restrictions with the ICNIRP values. These limits also agree with those defined by the European Commission in [125].

In table 4.1 two exposure situations are defined. The occupationally exposed population consists of adults who are generally exposed under known conditions and are trained to be aware of potential risks and to take appropriate precautions. By contrast, the general public comprises individuals of all ages and of varying health status, including particularly susceptible groups or individuals, taking into account that in many cases members are unaware of their exposure to electromagnetic fields. Moreover, members of the general public can not reasonably be expected to take precautions to minimize or avoid exposure, thus more stringent restrictions are adopted for the general public than for the occupationally exposed population [126].

Finally, it is important to point out that the aforementioned documents are the most important recommendations and guidelines, however other documents which give specific details can be found. For instance [130] shows several measurement methods to evaluate the SAR inside the human body, [131] presents an extensive review of experimental studies about the biological effects of RF exposure and [132] gives information about the measures employed by different countries. In this last document it is stated that the Spanish law [133] includes the contents of the European Commission recommendation of July, 1999 [125].

4.1.2 Antennas implanted inside the human body

This section will analyze the scenario where one of the antennas is implanted inside the body and a wireless link with an external antenna is desired. The implanted antenna (with a thin insulation layer of biocompatible material) will be in contact with the human tissues, critically affecting the antenna performance. For this reason the antenna must be designed taking into account the conditions that it will find inside the body.

Another sensitive issue is signal propagation through the body. Body tissues are highly lossy, especially at high frequencies, therefore, they will introduce additional losses degrading the implanted antenna gain and the wireless link performance. Most

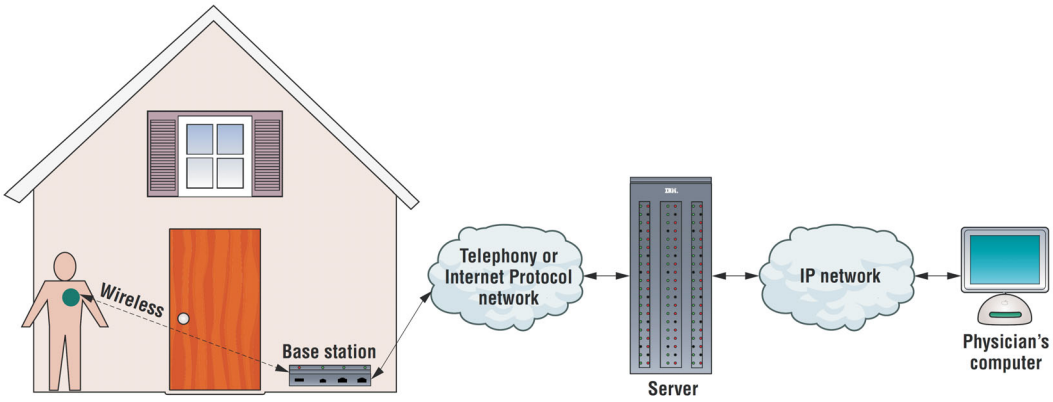


Figure 4.1: *Recent implantable cardiac defibrillators provide in-home monitoring via wireless base stations that relay data to doctors [136].*

of the efforts of this chapter will be focused on the analysis and quantification of these propagation losses. In spite of all these drawbacks, implantable antennas are an important field of interest that is growing rapidly.

During the second half of the past century an important effort was put in the design of implantable biomedical devices for the treatment of various illnesses [134]. In 1952, Dr. P. Zoll reported the first external pacemaker [135], and some years later, the first implantable pacemaker was presented. As time went by, these primitive devices evolved to the current ones, with better performance and longer lifetime. Other examples of applications with implantable devices are monitoring systems, drug delivery systems and defibrillators [134].

In these implantable devices a wireless link with an external receiver can be used to monitor the device battery level and the clinical parameters of the patient. As an example of application, an implantable sensor with a wireless communication link could be used in a glucose monitoring system [137], automatically alerting the emergency systems when a problem is detected (figure 4.1). These implantable sensors with wireless links will be one of the key points for future e-health systems, especially for the treatment of chronic diseases. One of the most important companies in this area is Medtronic [20], a company founded more than 40 years ago that offers a wide catalog of solutions for the treatment of different illnesses (figure 4.2).

The most commonly used frequency band for implantable antennas is the Medical Implant Communications Service (MICS) band allocated in 1999 by the FCC in the $402 - 405$ MHz frequency range. It is a frequency band for ultra-low power, unlicensed, mobile radio services for transmitting data in diagnostic or therapeutic applications

(a) (b)

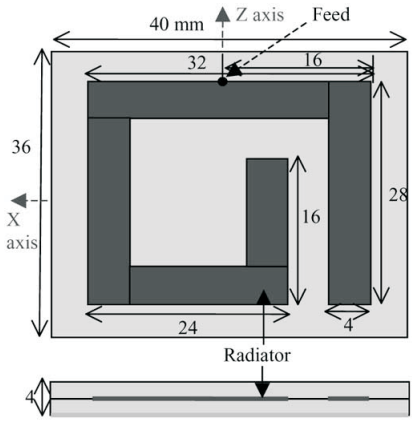
(c)

Figure 4.2: *Medtronic offers a wide catalog of solutions for the treatment of chronic diseases [20]. (a) Real time glucose monitoring system. (b) Implantable cardioverter defibrillator with wireless telemetry system. (c) Implantable cardiac monitor.*

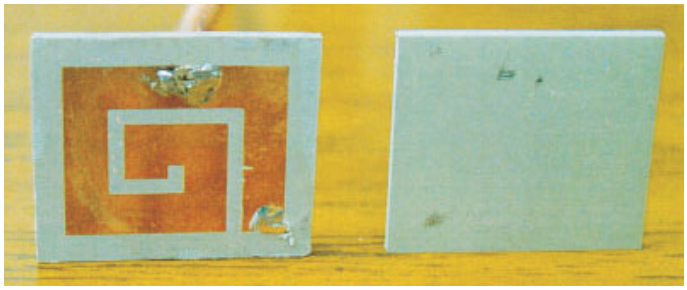
associated with implantable medical devices. Another option is the use of low frequency (13.56 MHz) inductive links [103, 138], but low ranges are obtained (4 cm) and the external coil must touch the patient body near the internal coil [103].

In [139] and [140] two antenna designs for the MICS band with a superstrate layer to improve the antenna performance were presented (figure 4.3). Due to their shape and dimensions (4 cm x 3.6 cm x 0.4 cm for the microstrip antenna and 2.4 cm x 2 cm x 0.25 cm for the PIFA), these antennas are suitable for big medical devices implanted in the human chest, like pacemakers or defibrillators. In [141] a similar design is studied, analyzing the effects of the pacemaker and battery pack on the antenna performance. Other examples of antennas in the MICS band can be found in [142–147].

The aforementioned examples, and the existing commercial solutions, typically use the MICS bands, however there exists a growing interest in higher frequencies, such as the 2.45 GHz ISM band, because smaller antennas can be obtained. For instance in [104] an implantable antenna for an intracranial pressure monitoring system (figure 4.4(a)) was presented. [148] and [149] analyze the cavity slot of figure 4.4(b) and [150] shows an implantable wireless telemetry board with an ISM band antenna for invivo transocular transmissions.



(a)



(b)

Figure 4.3: *Antennas for implantable devices. (a) Microstrip antenna. [139]. (b) PIFA antenna. [140].*

In [144, 151] a dual band PIFA antenna for the MICS and the 2.45 GHz ISM bands is shown (figure 4.4(c)). This dual band design can be used with the ultra low power implantable transceiver Zarlink ZL70101 [152] that transmits information in the MICS band and uses a wake-up signal at 2.45 GHz. Finally, of interest is the study presented in [153], where the radiation characteristics of an ingestible wireless device in different frequency bands are analyzed.

Guidelines and regulations for antennas implanted inside the human body

As in the first scenario, or even more because here the antennas are inside the body, the safety issues related to the effects of electromagnetic fields on the body tissues are a point of interest and awareness.

As was previously said, the most commonly used frequency bands, especially in commercial applications, is the MICS band. It is a free band for implantable devices that can be used without a specific license. Nevertheless, according to [19], the maximum

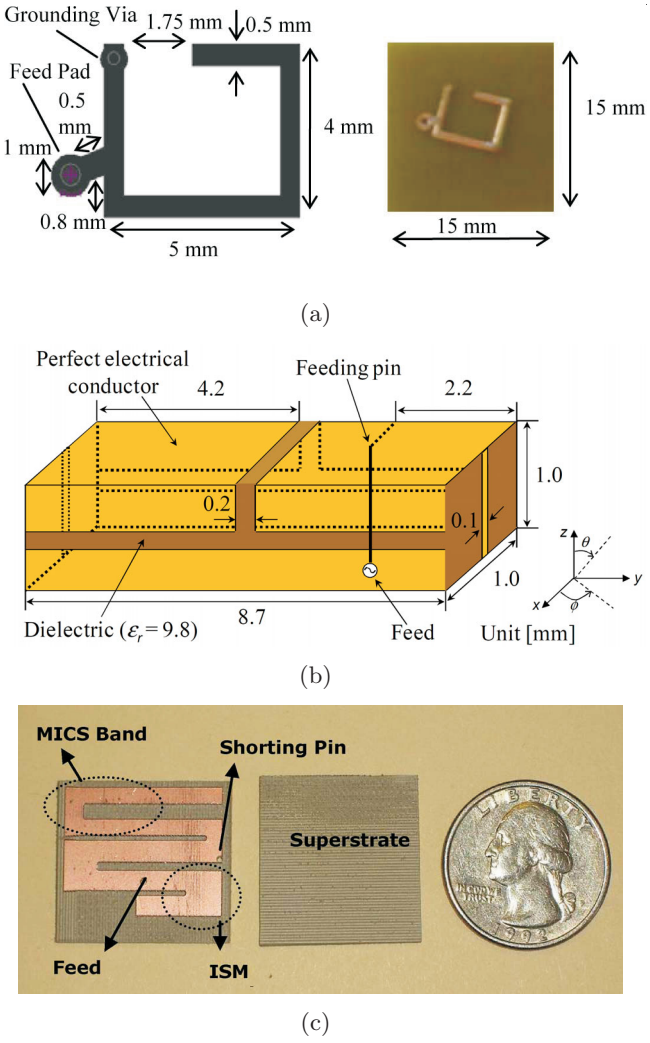


Figure 4.4: *Examples of antennas operating in the 2.45 GHz ISM band. (a) PIFA antenna for an intracranial pressure monitoring system [104]. (b) Cavity slot antenna to be implanted in the patient arm [148]. (c) Dual band antenna for the MICS and ISM bands [144].*

power in any 300 KHz bandwidth radiated by medical radio transmitters operating at 402 - 405 MHz shall not exceed 25 μw EIRP. The FCC also specifies how the EIRP measurements must be carried out. The radiated emissions and EIRP levels may be determined by measuring the radiated field from the equipment under test at a distance of 3 meters [19]. Furthermore radiated emissions and EIRP measurements must be made in accordance with a commission-approved human body simulator and test techniques. A suitable tissue substitute material is defined in the OET Bulletin 65 Supplement C [154].

Soon afterwards the definition of the MICS band, some regulatory arrangements

supporting MICS devices were done in order to enable the use of implantable devices worldwide. For instance the International Telecommunication Union Radiocommunications sector (ITU-R) developed a recommendation for the MICS band [155] taking into account that this band has to be shared with the Meteorological Aids Service (allocated in the same frequency range some time ago). However, it is not a problem because with the EIRP limit of MICS systems and the interference mitigation techniques used by the MICS equipment, no harmful interference should occur between both systems [155].

The European Radiocommunication Commission also allocated the same band for the same services in the recommendation ERC/REC70-03 [156]. [156] covers all kind of Short Range Devices (SDR), but annex 12 is focused on medical implants, under the name of Ultra Low Power Active Medical Implants (ULP-AMI). In this recommendation the maximum EIRP power of 25 μw defined by the FCC [19] is also used. The European Telecommunications Standards Institute (ETSI) developed harmonized standards for these devices [157] where some additional information about the measurements methods to test the ULP-AMI limits is given.

The aforementioned standards and recommendations are for the MICS band ($402 - 405$ MHz), but in recent years an important effort has been put into the design of antennas for implantable devices working at higher frequencies. Some examples that use the 2.45 GHz ISM band have been shown in figure 4.4. To the author's knowledge, there is not a well defined specific regulation for implantable devices at 2.45 GHz. For this reason, generally the SAR limits given for the first scenario where the antennas were in the vicinity of the human body are also used when one of the antennas is inside the body. However, in implantable sensors the transmitted power levels are usually very low in order to increase the battery life, and hence the SAR levels are below these limits. Some other authors like K. Ito in [148], also use the power limit of 25 μw for the 2.45 GHz ISM band.

4.2 Human body models

Implanted antennas performance evaluation is difficult. In simulations accurate models of the human body are necessary. In measurements, physical models called phantoms filled with special liquids are normally used. These liquids have electrical properties similar to body tissues and allow the evaluation of the antenna and the whole system performance under test conditions similar to the real body. In this section, different solutions for modeling the body effects will be analyzed.

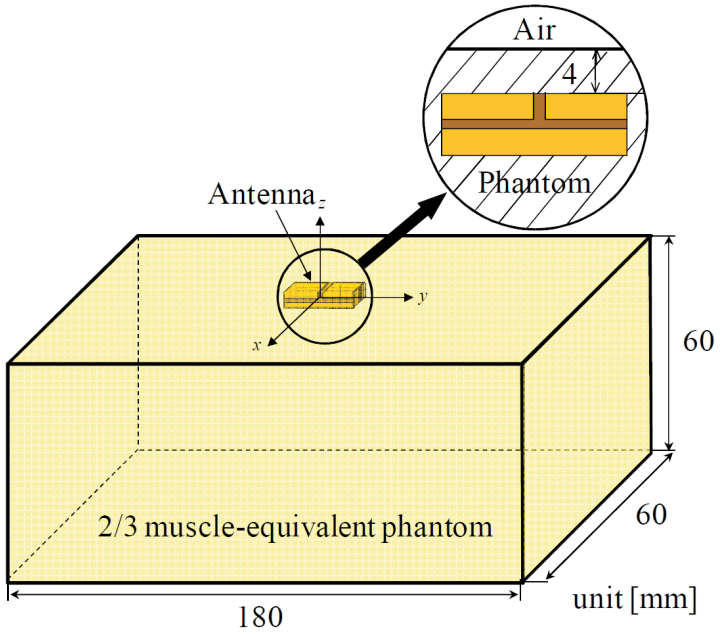


Figure 4.5: *One layer model used by K. Ito for the antenna simulation inside the human body [148].*

4.2.1 Simulations with human body models

In simulations, the simplest and fastest, but least accurate option is the use of a single layer structure of an homogenous material with properties similar to body tissues. As can be seen in figure 4.5, the antenna is simulated inside a block of this dielectric material [140, 141, 148]. However, in the one layer model there is an additional question that must be answered: What are the properties of the equivalent material used to model the overall behavior of the complex human body dielectric structure? In literature multiple solutions can be found. For instance, [149] and [158] used a material called "2/3 muscle", that is a material with the electrical properties of the muscle tissue multiplied by 2/3.

In spite of the fact that "2/3 muscle" material has been used by several authors, the Office of Engineering and Technology (OET) of the FCC published some guidelines for the SAR evaluation process (OET Bulletin 65 supplement C [154]). This document gives the properties of the equivalent material that must be used (table 4.2). The conductivity of this equivalent material is slightly higher than the conductivity of muscle and other body tissues, therefore the propagation losses simulated with this equivalent material will be slightly overestimated.

Another possible option is the use of a multilayer structure of finite or infinite dimensions. Normally, a three layer (muscle - fat - skin) model is used [139, 159, 160].

Table 4.2: *Equivalent tissue dielectric parameters for head and body models [154].*

	Head		Body	
Frequency (MHz)	ε_r	$\sigma(S/m)$	ε_r	$\sigma(S/m)$
150	52.3	0.76	61.9	0.80
300	45.3	0.87	58.2	0.92
450	43.5	0.87	56.7	0.94
915	41.5	0.98	55.0	1.06
1450	40.5	1.20	54.0	1.30
1610	40.3	1.29	53.8	1.40
1800 - 2000	40.0	1.40	53.3	1.52
2450	39.2	1.80	52.7	1.95
3000	38.5	2.40	52.0	2.73
5800	35.3	5.27	48.2	6.00

Table 4.3: *Electrical properties of body tissues in different ISM frequency bands [161].*

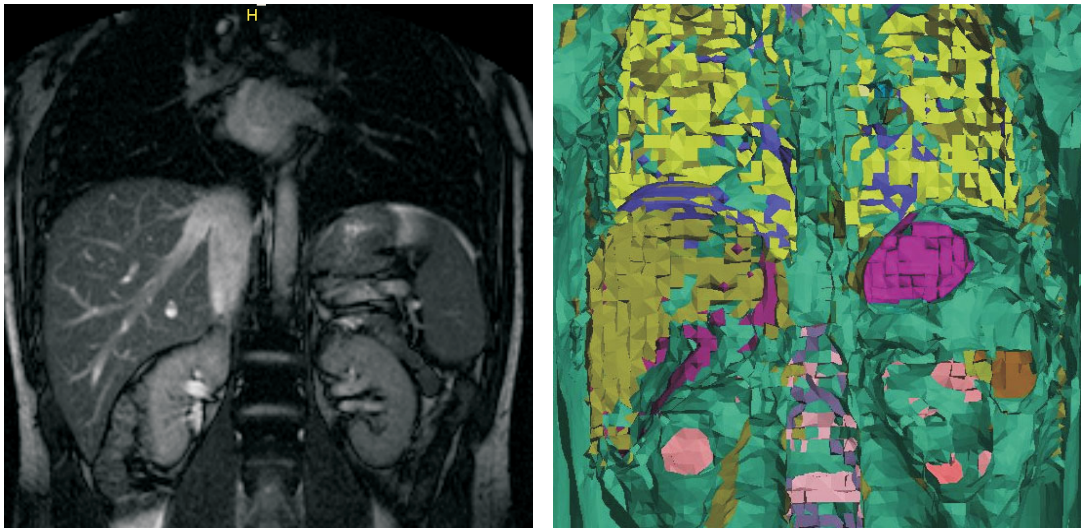
	Relative Permittivity ε_r				Conductivity σ (S/m)			
Frequency (MHz)	433	915	2450	5800	433	915	2450	5800
Skin	46.08	41.33	38.01	35.11	0.70	0.87	1.46	3.72
Fat	5.57	5.46	5.28	4.95	0.04	0.05	0.10	0.29
Muscle	56.87	54.99	52.73	48.48	0.80	0.95	1.74	4.96

The electrical properties of body tissues can be found in [161], and are summarized in table 4.3, where the skin, fat and muscle electrical parameters in some ISM bands are shown. The human head also can be modelled wit a multilayer model, like in [139] where a spherically six layer (brain - cerebrospinal fluid - dura - bone - fat - skin) head model is used.

If the antenna is outside the body, similar results are obtained with the one layer and the three layer models. However, as it will be shown later in this chapter, a three layer model is necessary when the antenna is inside the body because the antenna behavior can differ depending on the tissue layer where it is implanted.

Finally, there is the option of using 3D anatomical models [153, 162–165]. These realistic models offer the most accurate results because almost all the details of the body shape are taken into account. There are two kinds of anatomical models: the homogeneous models and the inhomogeneous models. In the first ones, the model consist of a uniform dielectric material block with the shape of a human being, meanwhile in the second ones the different body tissues are taken into account.

Inhomogeneous models can be created with the information of Magnetic-Resonance Imaging (MRI) [164] images where the shape of the internal organs can be obtained.



(a) (b)

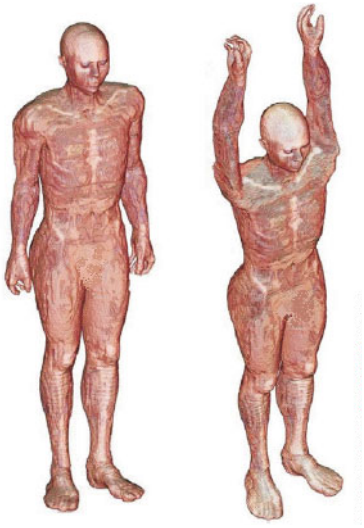
Figure 4.6: *The human body is an inhomogeneous structure. (a) MRI image of the thesis author. (b) Cut of the FEKO inhomogeneous body model. The simulation model takes into account most real body details.*

The body tissues properties at different frequencies appear in [161]. Figure 4.6(a) shows a MRI image of the thesis author and figure 4.6(b) shows a cross section of a FEKO inhomogeneous model. Comparing both figures, it can be seen that the inhomogeneous model takes into account most real body details.

Furthermore, models with different positions and kinds of person (male, female, adult, child, ...) can be used to evaluate the effects of different body parameters in the system performance. Many examples of human body models can be found in literature (figure 4.7) and some commercial softwares, like FEKO [70], Semcad X [166] or Empire XCcel [167], also offer their own models.

In particular, the FEKO software used in this thesis provides several models of specific parts of the body, like a model of the head and the shoulders for mobile phones SAR evaluation, and three whole body models: one with a superficial mesh (SEP model), and two with a volumetric mesh to be solved with Finite Elements Methods (FEM models). The SEP model is a homogenous anatomical model. One of the FEM models is homogenous (334,733 tetrahedra) whereas the other is inhomogeneous (2.2 millions tetrahedra). Both are suitable up to 1 GHz [70].

Table 4.4 shows the peak memory usage and the simulation time necessary for the simulation of a 900 MHz half wavelength dipole antenna in front of a human body. When the inhomogeneous model is used computational requirements are high, the simulation



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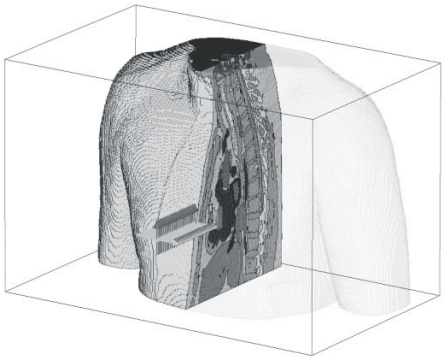

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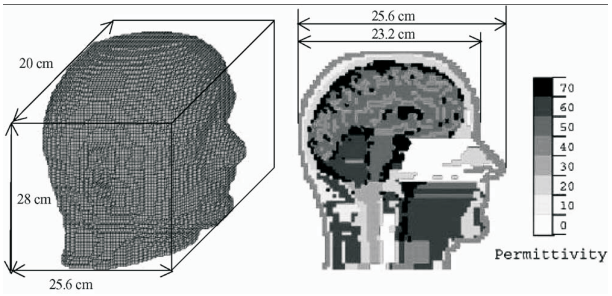


(c)

Figure 4.7: *Examples of human body models found in literature. (a) 3D human body in different positions [164]. (b) Cut of an inhomogeneous body model [168]. (c) 3D inhomogeneous head model [139].*

Table 4.4: *Computational requirements of FEKO body models.*

Model	Peak Memory (GB)	Simulation time (s)
Homogeneous SEP model	9.45	7669
Homogeneous FEM model	2.61	1678
Inhomogeneous FEM model	15.62	33618

needs 15.6 GB of memory and the analysis of each frequency takes 33600 seconds. With the homogeneous models the requirements are smaller, for instance the FEM homogeneous model only needs 2.6 GB of memory and the simulation time is below 1700 seconds. These results show that generally the simulations made with homogeneous models are faster and need less memory than the simulations made with inhomogeneous models. Therefore there

exists a clear trade-off between the simulations accuracy (inhomogeneous models) and the simulations requirements (homogeneous models).

Furthermore, it is important to take into account that in some scenarios homogeneous models can be perfectly valid, for instance, when a dipole antenna is placed in front of a human body (figure 4.8). In this kind of simulation where the antenna is external to the body, homogenous and inhomogeneous models give the same results. Nevertheless, when the antenna is implanted inside the body normally an inhomogeneous model has to be used to obtain accurate results.

The main problem of 3D anatomical models is that computational requirements are high due to the models complexity and the need for a fine mesh to take into account all the body details. For this reason, in spite of the better accuracy of the anatomical models, the aforementioned one layer and three layers models are still very popular.

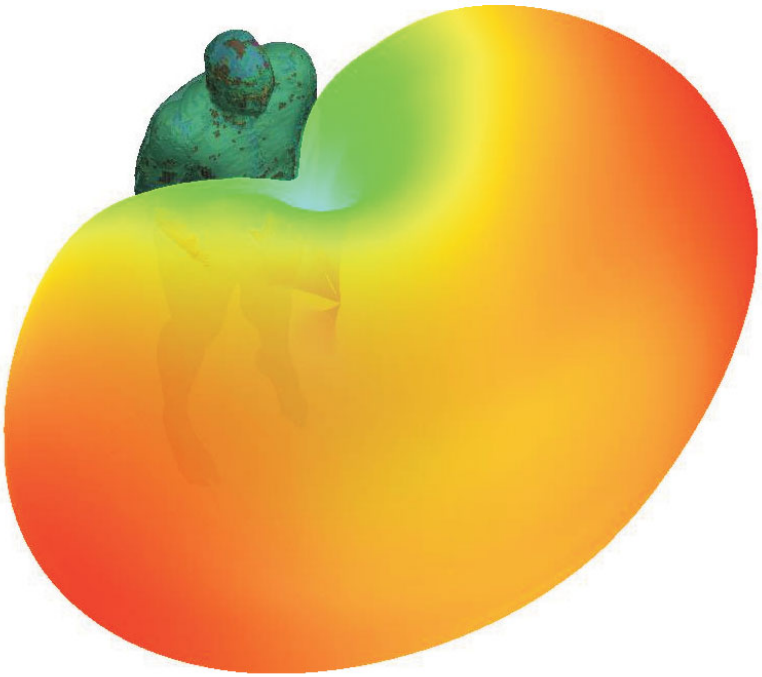


Figure 4.8: *Radiation pattern of a dipole antenna in front of a human body at a distance of 10 cm. The inhomogeneous FEM model was used, but the same results were obtained with the homogeneous one. A 30 dB dynamic range has been used for the 3D far field plot.*

4.2.2 Measurements using human body phantoms

Implanted antennas performance evaluation is not a trivial issue because it is not possible to carry out measurements in the real operating scenario, so a scenario that emulates the environment of the real human body is necessary.

The most widespread solution is the use of a container with a regular or an anatomical shape filled with an special liquid [169]. The antenna is sunk into the liquid and measurements are carried out [139, 140, 148, 149, 158]. These liquids, also known as body tissues simulating liquids, must have dielectric properties similar to the human body. There are some commercial liquids [166] with the dielectric properties of table 4.2 ready to carry out this kind measurements, and SPEAG also offers complete solutions to automatically carry out these measurements, like the Dasy 5 measurement setup shown in figure 4.9 [166].

If a container with a regular shape is used (for instance a cuboid), measurements are equivalent to the one layer model simulations, meanwhile if the container has an anatomical shape, measurements are equivalent to simulations with an homogeneous anatomical body

Figure 4.9: *Dasy 5 measurement setup. It is designed for SAR measurements, but it could be also used for the measurement of implanted antennas [166].*

model. In section 4.4 some measurements of a patch antenna using body tissue simulating liquids with a regular shape container will be presented to demonstrate the validity of the simulator predictions.

Recently Karacolak [144, 151] has been working on the use of tissue simulating gels with dielectric properties similar to the body tissues. This is an option of interest because it is a first attempt to measure using multilayer structures. There are also some authors that have carried out measurements with real tissues, but only those of animals [150, 170].

4.3 Study of the propagation losses introduced by body tissues

One of the main problems related to implanted antennas are the additional propagation losses introduced by the body tissues that critically degrades system performance. This section will focus on the analysis of these losses.

The propagation losses of body implanted antennas in the ISM bands of 433 MHz, 915 MHz, 2450 MHz and 5800 MHz will be studied. The results of the 433 MHz band can be also considered valid for the 402 - 405 MHz MICS band. It is important to point out that the aim of this study is the analysis of the body effects on propagation losses, not the design of the implanted antenna, for this reason simple straight dipoles will be used as transmitting and receiving antennas. Straight dipoles may not be the optimal solution, but they are typically used as reference antennas.

Two body models will be considered, one based on a single equivalent layer and the other based on a three layer structure, showing the advantages and limitations of each one. The study with anatomical models will not be carried out due to the high system requirements and simulation times. Firstly, the one layer and three layer models will be used to evaluate the effects of introducing the antenna inside the body. The system performance as a function of several parameters like the external antenna distance, implanted antenna depth and operating frequency will be evaluated. Then, the effects of the different tissues thickness will be studied using the three layer model, and finally the consequences of adding a coating layer to the antenna, necessary for biocompatibility issues, will be discussed.

4.3.1 Free space wireless link

One of the main parameters that defines the performance of a wireless system is the

antenna pair gain (Ga) [75]. It can be calculated using the scattering parameters as

$$Ga = \frac{|S_{21}|^2}{(1-|S_{11}|^2)(1-|S_{22}|^2)} \tag{4.4}$$

where the term in the denominator is used to subtract the mismatch losses of the transmitting and receiving antennas.

The antenna pair gain is mainly related to the propagation losses and the antennas gain. In a free space scenario, where both antennas are outside the body and far field conditions are assumed, the antenna pair gain can be computed as

$$Ga_{(fs)} = G_T \cdot G_{Air} \cdot G_R \tag{4.5}$$

where G_T is the transmitting antenna gain, G_R is the receiving antenna gain and G_{Air} is the air propagation gain or free space propagation gain defined as

$$G_{Air} = \left(\frac{\lambda}{4\pi r}\right)^2 \tag{4.6}$$

where λ is the free space wavelength and r is the distance between both antennas.

4.3.2 One layer model

Now, the effects of having one of the antennas implanted inside the body will be studied using the one layer model. Firstly, the simulation scenario will be presented and then the simulated results obtained with the FEKO MOM code [70] will be analyzed.

Simulation scenario

The system under study is based on two dipole antennas, one implanted inside the human body, and one outside the body. Half wavelength resonant dipoles are used as transmitting and receiving antennas to avoid precision errors when subtracting the mismatch losses in equation 4.4. The simulation scenario and the definition of the different parameters are shown in figure 4.10. In order to reduce the simulation computational requirements an infinite dielectric structure was used to model the body, however the results were also checked with finite dielectrics. The equivalent tissue properties can be found in table 4.2.

Simulation results

The insertion of the antenna inside the human body has two main effects: A decrease in the resonant antenna dimensions and a degradation of the gain due to the high permittivity and conductivity of body tissues.

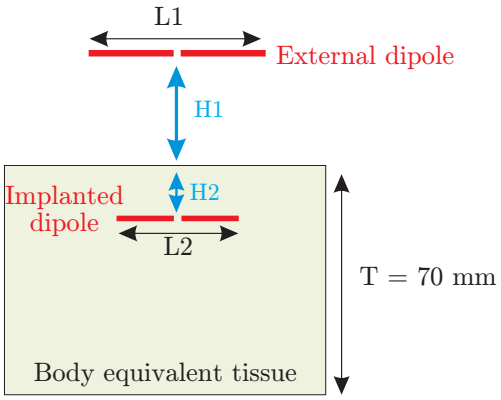


Figure 4.10: *One layer scenario. The transmitting and receiving antennas are half wavelength resonant dipoles. The equivalent tissue properties are shown in table 4.2.*

Figure 4.11(a) plots the antenna pair gain as a function of the internal antenna depth when the external antenna is 400 mm away from the body surface. The antenna pair gain decreases when the frequency is increased. For instance, if the implanted antenna depth is equal to 10 mm, the antenna pair gain decreases almost 20 dB when the frequency is changed from 2.45 GHz to 5.8 GHz. This degradation is due to the increase in the free space propagation losses (the frequency is multiplied by 2, therefore losses are increased in 6 dB) and the increase in the body conductivity (table 4.3) that introduces approximately 14 dB of additional losses.

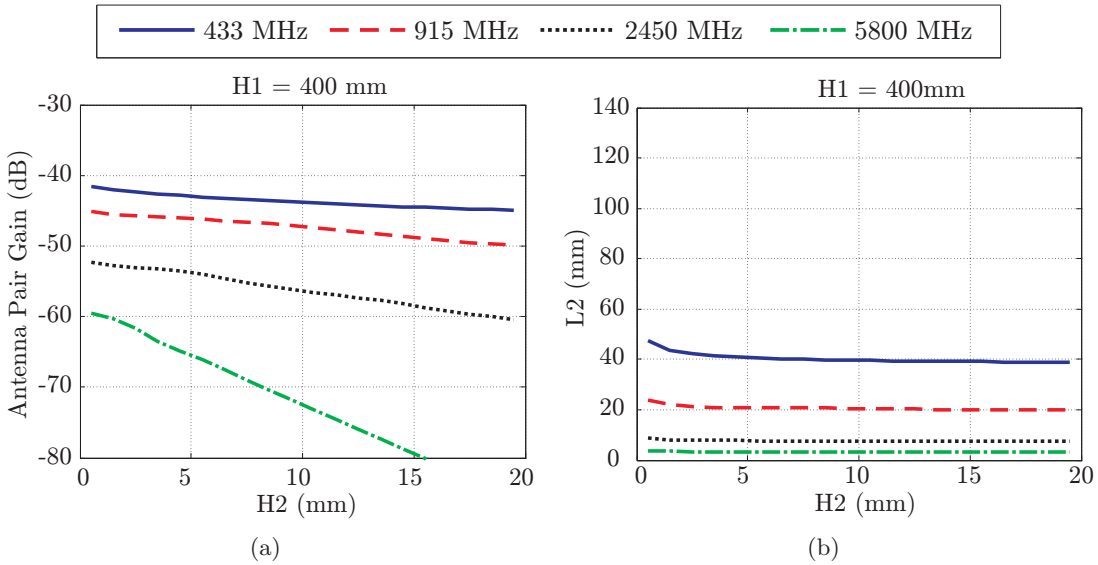


Figure 4.11: *Simulation results for the one layer model (figure 4.10). (a) Antenna pair gain. (b) Implanted dipole length to keep the antenna resonant at the desired frequency.*

Figure 4.11(b) shows the required implanted dipole length to keep the structure resonant at different depths. It was obtained empirically through an iterative simulation process. This length is inversely proportional to the frequency and it is significantly smaller than one half of the free space wavelength because of the high permittivity of body tissues. With the one layer model the dipole length does not significantly change with depth (figure 4.11(b)).

According to these results, a clear trade-off between antenna dimensions and antenna performance exists. When frequency is increased, antenna dimensions are reduced, but the antenna pair gain decreases degrading the system performance.

4.3.3 Three layer model

In this section we will study the propagation losses using a three layer model to simulate more accurately the real scenario and we will highlight the one layer model limitations.

Simulation scenario

In this scenario the body effects were modeled using a three layer model made of skin, fat and muscle (figure 4.12). Depending on the body area where the antenna is implanted the thickness of each layer can be different. For this reason thickness values used in literature may vary from author to author [137, 139, 144, 171, 172]. We used the thickness shown in figure 4.12. The body tissues properties can be found in table 4.3.

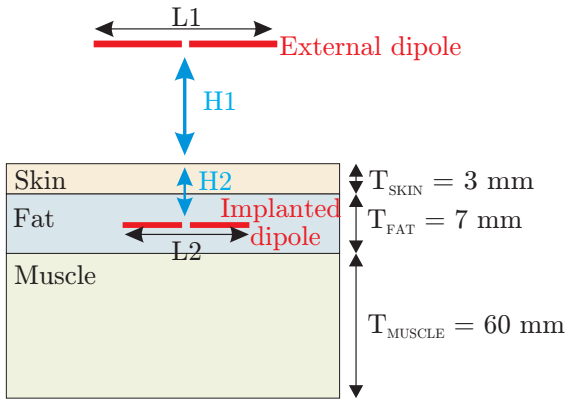


Figure 4.12: *Three layer (skin - fat - muscle) model simulation scenario. As in the previous scenario, the transmitting and receiving antennas are half wavelength resonant dipoles. The body tissues properties are shown in table 4.3.*

Simulation results

Figure 4.13 shows the simulation results with the three layer model. The antenna pair gain (figure 4.13(a)) significantly depends on the layer where the antenna is positioned. In the implanted antenna gain plot (figure 4.13(b)), the effects of the three layers can be clearly distinguished. According to table 4.3, skin and muscle tissues have higher conductivity values than fat, therefore the implanted antenna gain is lower in these layers. It must be noted that for the higher frequency bands, where the fat layer is electrically thicker, more than 10 dB of difference can be observed.

The air propagation gain (G_{air}) is shown in figures 4.13(c) and 4.13(d). Simulated air propagation gain has been obtained subtracting the internal and external antennas gain from the antenna pair gain (Ga). The theoretical results using a free space propagation model (equation 4.6) are plotted with a yellow line. When the implanted antenna depth (H_2) is changed, flat plots are obtained, proving that with typical antenna depths of some millimeters, the air propagation gain does not significantly depend on this parameter. When the distance to the external antenna (H_1) is changed, an inversely proportional quadratic dependence is observed, showing a good agreement with the theoretical model. These results prove that when far field conditions are fulfilled, body losses can be included in the implanted antenna gain and use a simple free space propagation model to evaluate the system link budget.

When the implanted dipole is in the fat layer, the effective permittivity is significantly lower than in the other layers and the dipole length (L_2) has to be increased by approximately a factor of 2 to keep the same resonance frequency (figure 4.13(e)) Again, we can see the aforementioned trade-off between antenna dimensions and performance. When the antenna is in the fat layer, its gain is higher (figure 4.13(b)), but it has to be longer (figure 4.13(e)).

In the coverage maps of figure 4.14, the antenna pair gain is plotted as a function of the implanted antenna depth (H_2) and the distance to the external receiver (H_1). These maps can be useful in the system design process. From the system specifications, the required antenna pair gain is calculated, and then the system range for a given frequency band and an implanted antenna depth can be easily predicted with the information of figure 4.14. As an example, if the antenna pair gain has to be higher than -35 dB to assure the proper system operation, we can predict that it will be impossible to have the system operating at 5.8 GHz.

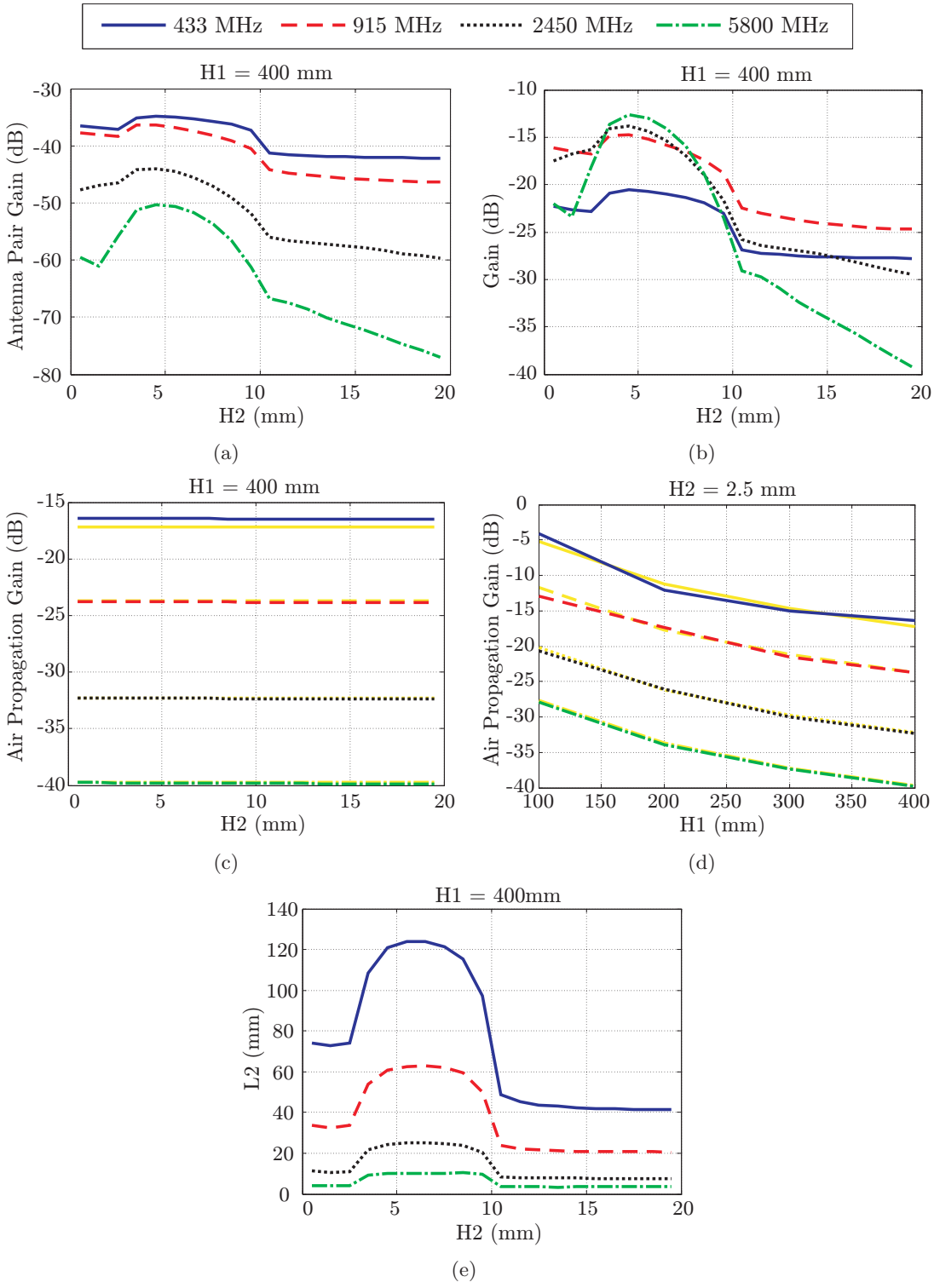


Figure 4.13: *Simulation results with the three layer model (figure 4.12). Yellow lines show the theoretical results. (a) Antenna pair gain. (b) Implanted antenna gain. (c) Air propagation gain as a function of the implanted antenna depth. (d) Air propagation gain as a function of the distance to the external antenna. (e) Implanted dipole length to keep the antenna resonant at the desired frequency.*

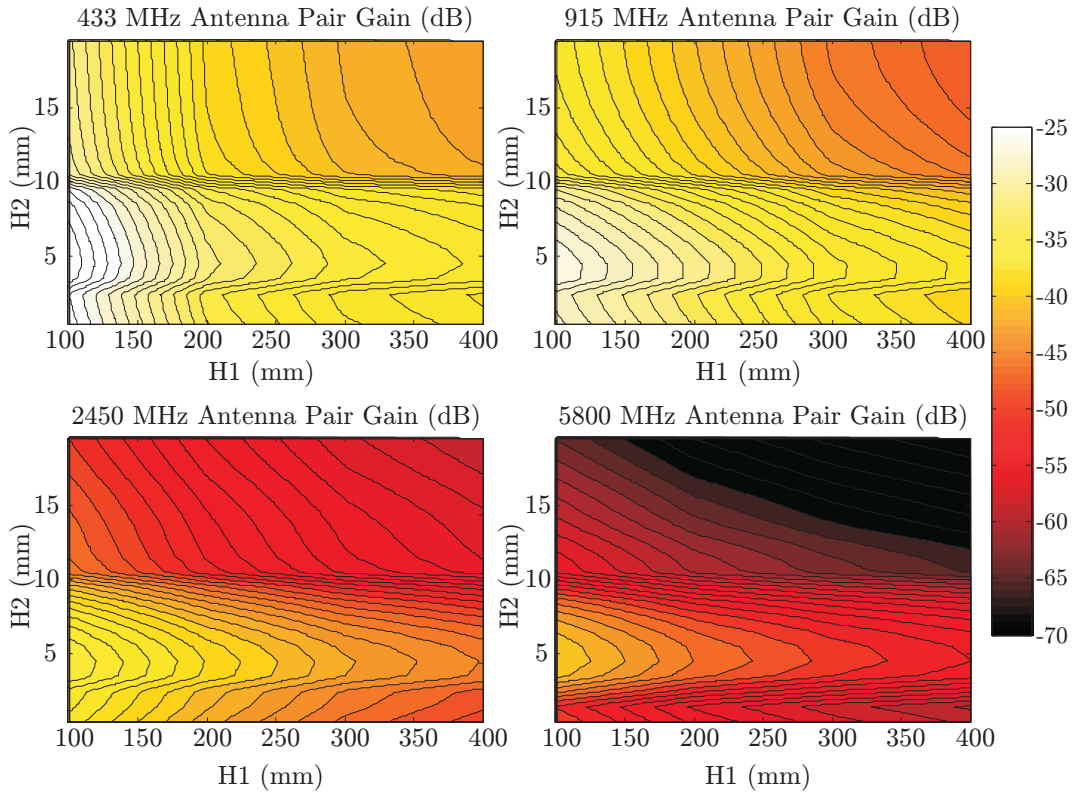


Figure 4.14: *Coverage maps showing the antenna pair gain (Ga) at different frequencies as a function of the implanted antenna depth and the distance to the external antenna. Coverage maps are useful to predict the system performance as a function of the frequency band and the antennas position.*

One layer model vs. three layer model

This section shows a comparison between the results obtained with the one layer model and the three layer model. The objective is to point out the limitations of the one layer model, which are similar to the limitations existing in the measurements carried out using phantoms filled with an homogeneous body tissue simulating liquid.

With reference to the antenna pair gain, in the one layer model (figure 4.11(a)) it is lower than in the three layer model (figure 4.13(a)). This was expected since the conductivity values used in the one layer model are slightly higher than the values used in the three layer model. Furthermore, it must be noted that the one layer model does not take into account the gain improvement when the antenna is in the fat layer. It tends to overestimate losses, so it can be accepted as a conservative approach from the point of view of the antenna pair gain.

In the three layer model differences in implanted dipole length depending on the tissue

where the antenna is implanted are significant and can be greater than a factor of 2 (figure 4.13(e)). This effect is not considered by the one layer model where the dipole length is kept almost constant (figure 4.11(b)). This is an important limitation of the one layer model and the measurements made using body tissue simulating liquids. If the antenna design is carried out taking into account the one layer model, the phantom measurements can agree with simulations, however, if the antenna is implanted in the fat layer, it will be detuned and the system might not work properly. The conclusion is that a three layer model must be used in the implantable antenna design process, especially if the antenna has to be implanted in the fat layer.

4.3.4 Effects of the layers thickness

We will show the effects of modifying the thickness of the skin and fat layers. The skin layer can change from 1 mm to 5 mm depending on the body part, and the fat layer from 3 mm to 15 mm.

The thicker the fat layer, the wider the depth range where the gain is maximum and the higher the maximum value of this gain (figure 4.15(a)). This is due to the fact that the antenna is further separated from muscle and skin, so the losses introduced by these tissues are reduced. Furthermore, as the thickness of fat layer increases, the depth range

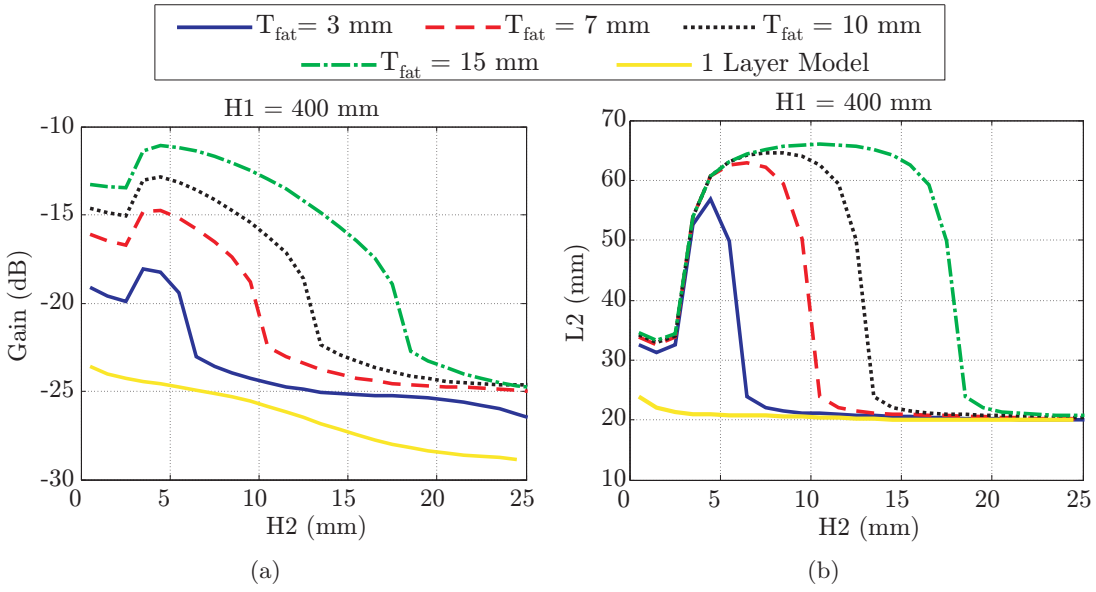


Figure 4.15: *Effects of the fat layer thickness. Yellow lines show the results obtained with the one layer model. Operating frequency $f = 915\,MHz$. (a)Implanted antenna gain. (b) Implanted dipole length to keep the antenna resonant at the desired frequency.*

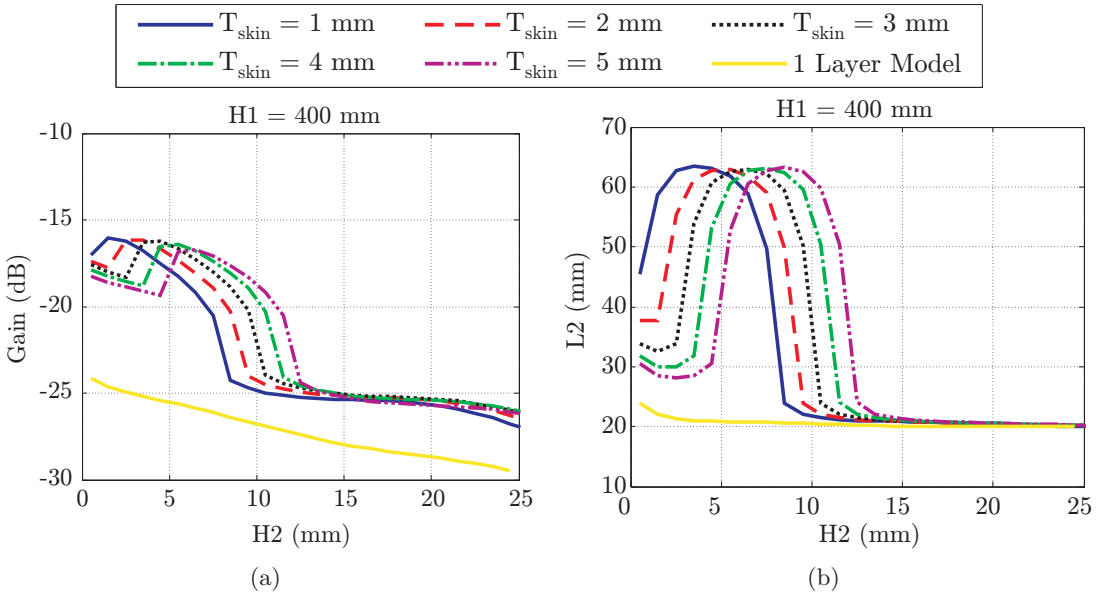


Figure 4.16: *Effects of the skin layer thickness. Yellow lines show the results obtained with the one layer model. Operating frequency $f = 915\,MHz$. (a) Implanted antenna gain. (b) Implanted dipole length to keep the antenna resonant at the desired frequency.*

where the dipole has to be longer is wider (figure 4.15(b)). Comparing the one layer model and the three layer model, the differences between both models become more significant when the fat layer is thicker.

When the skin layer thickness is increased the depth range where the antenna gain is higher changes accordingly to the fat layer position (figure 4.16(a)). The same conclusion can be drawn for the depth range where the antenna has to be longer (figure 4.16(b)).

4.3.5 Effects of a coating layer

In previous sections the antenna was directly in contact with the body. In real applications an insulation or coating layer of biocompatible material will be necessary to protect the antenna from the body tissues. Now, the effects of this coating layer will be analyzed.

In [173] Merli shows a list of materials that can be used as insulation layers. In our study, a low-loss coating layer with a relative permittivity equal to $\varepsilon_r = 4$ was used. Figure 4.17 shows the resonant dipole gain and length in the 2450 MHz ISM band when the coating layer thickness varies from 0.1 mm to 0.8 mm. Results with no coating are also plotted.

With regard to the antenna gain, the presence of a coating layer is advantageous,

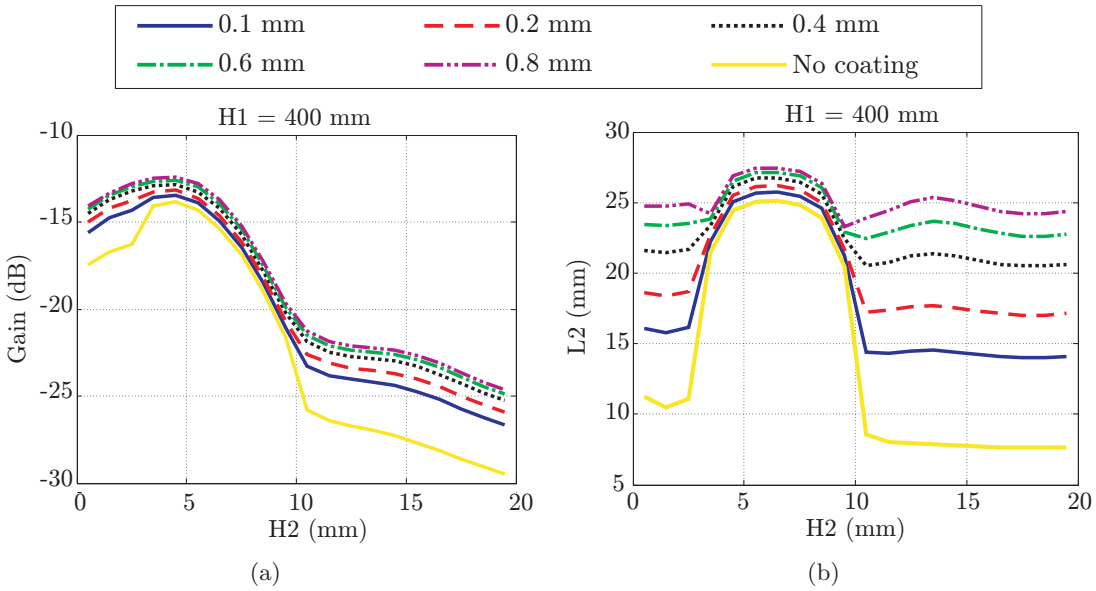


Figure 4.17: *Effects of a coating layer on the antenna performance. Operating frequency f = 2450 MHz. (a) Implanted antenna gain. (b) Implanted dipole length to keep the antenna resonant at the desired frequency.*

especially if the antenna is in the muscle or skin layers where gain improvement of around 5 dB can be obtained with coating layers thinner than 1 mm. When the antenna is in the fat layer, the gain improvement is not so significant because losses are not so great in this layer.

From the point of view of the implanted dipole length, the use of a coating layer has a positive and a negative aspect. The positive one is that the antenna becomes less sensitive to the surrounding material as the coating layer thickness increases, and therefore, the antenna detuning in the different tissues will be less significant. The negative aspect is that when the coating layer is applied the effective permittivity decreases and consequently the miniaturization obtained by the high relative permittivity of body tissues is less important. Variations in the dipole length higher than a factor of 2 can be observed depending on the coating layer thickness (figure 4.17(b)).

4.4 In-body antenna measurements with body tissue simulating liquids

All the previous results are based on FEKO [70] simulations using infinite dielectrics to model body effects, but FEKO results have not been validated with measurements. In this section some antenna measurements with an antenna immersed in various liquids will

be shown in order to demonstrate the validity of FEKO simulations when the antenna is inside a dielectric material.

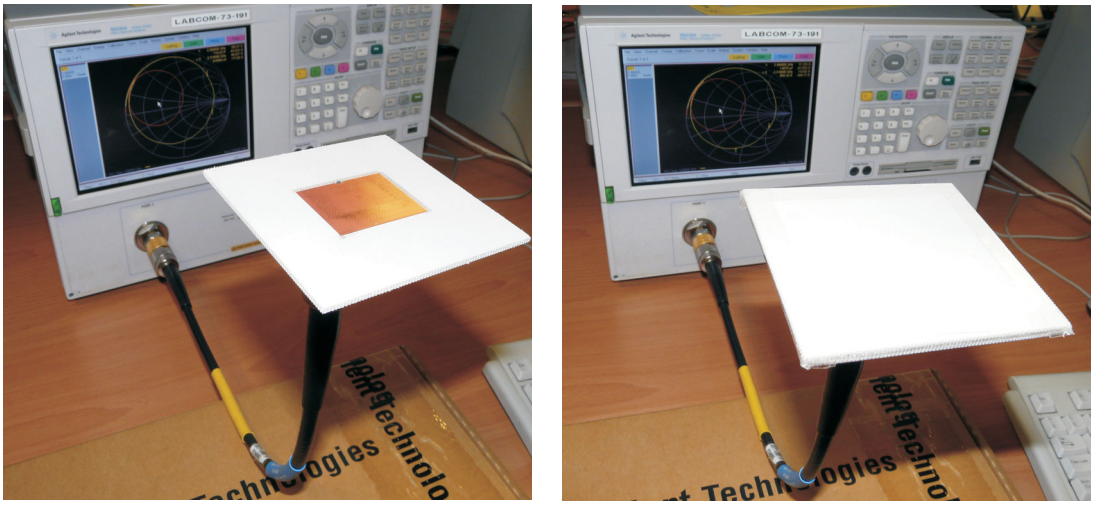
Measurements were carried out around the 2.45 GHz ISM band using a homogeneous liquid inside a rectangular receptacle, and hence they are equivalent to the one layer model used in the previous section. The antenna was measured immersed in two different liquids to show the simulations validity regardless of the conductivity and permittivity values of the liquid. First of all, measurements were made with distilled water ($\epsilon_r = 80$, $tan\delta = 0.1188$ at 2.45 GHz [174–176]), and then with the SPEAG body tissue simulating liquid MSL 2450 [166] ($\epsilon_r = 52.7$, $\sigma = 1.95$ S/m at 2.45 GHz). The MSL 2450 liquid conductivity and permittivity at 2.45 GHz are similar to human body dielectric properties.

The objective of this section is to show the validity of FEKO simulations when the antenna is inside a liquid (phantom), and not the design of the radiating element. Therefore a simple patch antenna with a superstrate was used in order to minimize the problems related to the measurement setup. In a patch antenna a differential feeding system is not necessary, avoiding the need for a balun and the possible problems related to unbalanced currents. Furthermore, the patch ground plane minimizes the undesired effects of the coaxial cable.

It is true that the study presented in section 4.3 was based on a dipole configuration and in this section a patch antenna is used to prove the FEKO simulations validity. However, if FEKO correctly models the body effects on a patch antenna, it is expected that it will also correctly model the same effects on a dipole. Moreover, it is important to point out that both kinds of antenna could be used in implantable devices. Patch configurations are useful for pacemakers, meanwhile a dipole antenna could be more suitable for a glucose sensor.

The microstrip patch was made on a Taconic TFR-45 [177] substrate ($\epsilon_r = 4.5$, $tan\delta = 0.0035$), and a superstrate of the same material was used to insulate the antenna from the liquid and improve its performance (figure 4.18). The patch and the ground plane dimensions were 28 mm x 35 mm and 80 mm x 80 mm respectively. The feeding point was at the edge of the patch to improve the matching levels when the antenna is immersed in liquid.

Figure 4.19 shows the measurement setup. The antenna under test was inside a rectangular plastic container filled with liquid. A circular plastic box with holes was used to hold the antenna at a fixed height. The antenna was fed using a SMA cable with a silicone gasket to avoid liquid leakages through the hole created to insert the cable into the container. One of the main advantages of this setup is that the antenna depth can be



(a) (b)

Figure 4.18: *Microstrip patch antenna with the feeding point at the edge of the patch to improve the matching levels when it is immersed in liquid. (a) Patch antenna without superstrate. (b) Patch antenna with superstrate.*

Figure 4.19: *Measurement setup. The antenna under test was immersed in liquid. The depth can be easily changed by adjusting the liquid level. A ruler was used to measure the antenna depth.*

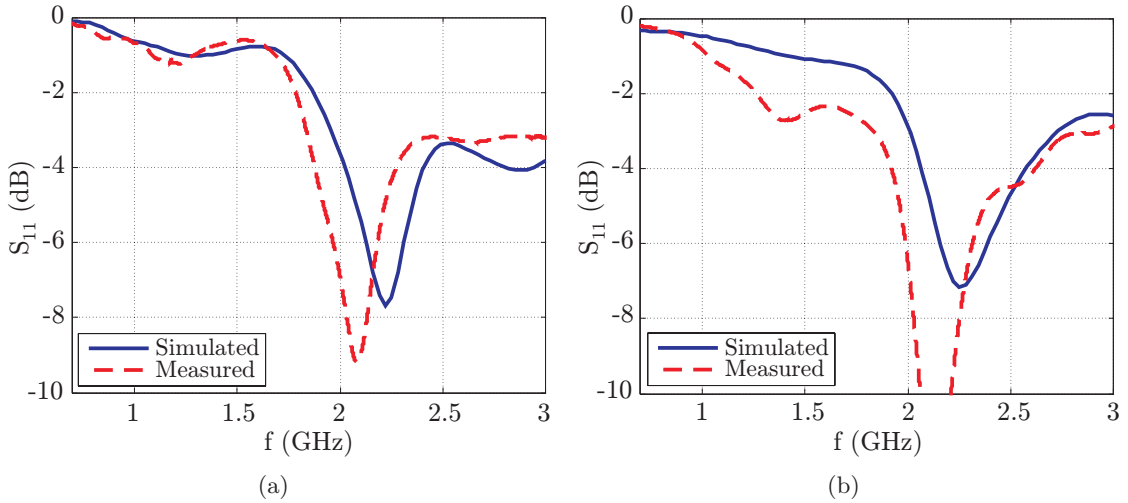


Figure 4.20: *Input reflection coefficient of a microstrip patch with superstrate (figure 4.18(b)) when the antenna depth is 8 mm. The measurement setup is shown in figure 4.19. (a) Antenna in distilled water. (b) Antenna in the SPEAG body tissue simulating liquid MSL 2450 [166].*

easily changed by adjusting the liquid level.

The input reflection coefficient of the microstrip patch was measured with the antenna immersed in distilled water (figure 4.20(a)) and in the SPEAG MSL 2450 liquid [166] (figure 4.20(b)). In both scenarios the input reflection coefficient is better than -6 dB, and good agreement between measurements and simulations is achieved demonstrating that the antenna input parameters predicted by FEKO when the antenna is immersed in a liquid are correct. Only a systematic frequency shift of less than 5 % is observed. This is mainly due to simulation inaccuracies and tolerances in the fabrication process and in the liquids permittivity and conductivity values. Furthermore, simulations do not take into account the frequency dependence of the liquids parameters.

Figure 4.20 demonstrates the validity of FEKO predictions of the input parameters. However in implantable antennas the most limiting parameter is usually the antenna gain. For this reason, measurements to prove the validity of FEKO predictions of this parameter were carried out. The antenna gain was obtained using the measurement setup shown in figure 4.21. The measured and simulated gain plots can be seen in figure 4.22.

The antenna depth was changed from 1 mm to 20 mm. Good agreement between simulations and measurements is achieved with both liquids. The general trend is that when the depth is increased, the gain is reduced because the electromagnetic waves have to propagate a higher distance through the lossy media. The gain in the MSL liquid is lower than in distilled water. This was expected because the MSL liquid conductivity is higher

Figure 4.21: *Measurement setup used to carry out propagation measurements. The receiving antenna was a microstrip patch with superstrate designed to operate in the desired frequency band (2.45 GHz ISM band) in free space. The distance between both antennas was 1.25 m. The "implanted" antenna depth was modified by changing the liquid level.*

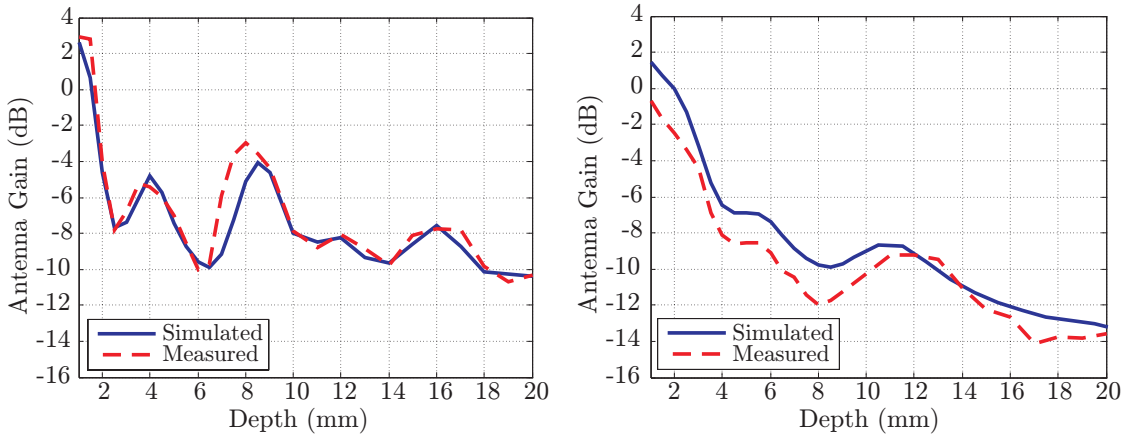


Figure 4.22: *Antenna gain in the normal direction of the microstrip patch with superstrate (figure 4.18(b)) as a function of the depth. The measurement frequency is near the 2.45 GHz ISM band. (a) Antenna in distilled water. (b) Antenna in the SPEAG body tissue simulating liquid MSL 2450 [166].*

than the water conductivity. Even with the MSL liquid, acceptable measured gain values near -10 dB are obtained. This is mainly due to the superstrate effects that significantly improve the antenna gain.

In this section we have compared the simulated and measured S_{11} parameter and gain when the antenna is immersed in two different liquids. Generally good agreement between simulations and measurements is achieved, demonstrating the validity of the FEKO simulation models used in section 4.3, and therefore, the validity of the propagation losses study.

4.5 Chapter conclusions

In some applications sensors are near or inside the body. The human body is a complex dielectric structure with high permittivity and conductivity values that significantly affects the antenna performance. In this chapter, these effects have been studied.

There are several regulations and guidelines that define the maximum allowed power levels for this kind of applications. In the few gigahertz frequency range, these maximum power levels are usually given as SAR limitations, 2 W/kg being the maximum peak SAR value permitted for the general public. There is a specific frequency band for implantable devices, the $402 - 405$ MHz MICS band with a maximum EIRP $= 25$ μw, however recent works use higher frequencies in order to reduce antenna dimensions.

The simulation of implantable antennas is complex because the body effects must be taken into account. In simulations the body effects are modeled using numerical models with properties similar to the human body called phantoms. Generally, the more detailed the simulation model, the more accurate the results but the higher the computational requirements. Furthermore, it is true that an inhomogeneous model can be necessary for implantable devices, but when the antenna is outside the body, a homogenous model is usually enough to obtain good results. In measurements, special liquids with electrical properties similar to the body tissues known as body tissue simulating liquids are used. These measurements are equivalent to the homogeneous simulation models.

A study to quantify the losses when an antenna is implanted inside the body has been presented in section 4.3. Simulations show that losses increase with frequency and antenna depth. Moreover, depending on the layer where the antenna is implanted its performance can be quite different. When the antenna is in the fat layer, its gain is better than when it is in the muscle or skin layers (with differences near 10 dB) but its length has to be longer to keep the same resonance frequency. The one layer model does not take into account these variations in antenna performance. This is the main limitation of the one layer model and of the measurements made using body tissue simulating liquids.

The effects of an insulation layer, necessary for biocompatibility issues, have been also analyzed. The presence of the coating layer is advantageous because it improves the antenna gain and the antenna becomes less sensitive to the surrounding materials properties. When the antenna is in the muscle or skin layers, gain improvements around 5 dB have been obtained with coating layers thinner than 1 mm. The negative aspect is that the antenna has to be larger because the miniaturization obtained by the high permittivity of body tissues is less important.

Finally some measurements of an antenna immersed in different liquids have been shown. FEKO simulations and measurements agree well, demonstrating the validity of FEKO predictions, and hence, the validity of the propagation losses study.

Chapter 5

Conclusions and future work

5.1 Conclusions

This thesis has been focused on the design of antennas for a distributed sensor network made up of remote sensors connected to a central unit with a wireless link. The central unit collects the sensors information and sends the data to the Internet with a WLAN connection that uses the 2.45 GHz and 5 GHz frequency bands defined by the 802.11 standards. The wireless link between the sensors and the central unit uses the 5.8 GHz ISM band, although it is also possible to use the 2.45 GHz ISM band to reduce propagation losses.

The sensor antenna and the central unit antenna are quite different. In the central unit a multiband antenna is necessary to give service to the 802.11 and the ISM bands with a single radiating element. Multiband functionality is not necessary in the sensors, but antenna dimensions are critical. The design objectives of both antennas are completely different, so for this reason they have been analyzed separately in two chapters.

Chapter 2 shows several multiband monopole configurations suitable for the central unit antenna. The designs presented in this chapter are based on two different concepts:

- Fractal geometries.
- Structures loaded with resonators.

Some fractal shapes, like the Sierpinski gasket, have been used in multiband antenna designs due to their self similarity property at different scale levels. Fractal shapes have been typically applied to the monopole element, but the results of this thesis show that

similar geometries can also be used in the ground plane to improve antenna performance. A fractal-based ground plane can reinforce the monopole resonances improving the existing bands, or even creating new operating bands. Two antenna designs have been presented, one where a triangular monopole is mounted on a fractal ground plane inspired by the Sierpinski gasket and one where a similar ground is used with a Sierpinski monopole. Both antennas fulfill the needs of the sensor network central unit, with good matching levels and similar radiation patterns in the different bands, therefore both solutions could be applied. Nevertheless, the design that uses the fractal shape in both elements is more flexible and it offers higher bandwidth in the lower band.

The second option is the use of resonator loaded structures. The multiband behavior of a monopole antenna can be improved loading the antenna with resonators. Furthermore, with the off-resonance loading principle the high losses of the resonators at their resonances can be avoided, introducing new bands without significantly degrading the in-band radiation efficiency. Based on these ideas several antenna designs have been presented.

The first is a dual band antenna based on capacitively loaded open loop resonators that covers the 2.45 GHz and 5 GHz bands of the 802.11 standard and the 5.8 GHz ISM band. The antenna operating principles have been analyzed, showing that the position of the bands can be easily controlled by adjusting the monopole and resonators dimensions. Next, two new designs that increase the number of operating bands have been shown. The first makes use of the SRR dual band behavior to achieve triple band performance. The second is an asymmetric structure that combines the resonances of two different resonators to obtain triple band performance. The third operating band available in these antennas could be used to improve the sensor network performance adding a WiMAX connection useful when the central unit is far from a WLAN access point.

All the solutions presented in chapter 2 are suitable for the central unit. If we have access to the ground plane design, the fractal-based solutions can be used, while if we do not have access to the ground plane design, then the solutions based on resonator loaded structures should be used. Moreover, both techniques could be combined to further improve the multiband performance using loaded monopoles on fractal-based ground planes.

Chapter 3 is focused on the remote sensors antenna design. A good option to obtain small and cheap sensors is the integration of the antenna and the electronic circuits into the same chip, obtaining a system-on-chip solution. To minimize costs, the antenna fabrication process must be compatible with the standard CMOS technologies. The main drawback is that the low resistivity silicon substrates typically used in CMOS technologies degrade the antenna performance.

Small sensors are desirable; therefore, miniaturization must be applied to reduce antenna dimensions. This chapter shows the performance offered by several miniaturization techniques when they are applied to antennas fabricated on silicon. The main problem is that the losses associated with miniaturization are emphasized by the use of a low resistivity substrate. For instance, the antenna gain decreases in 2.5 dB when the number of meanders is increased from 2 to 13.

Micromachining techniques are a good solution to reduce the silicon losses and improve antenna efficiency. The best option would be to eliminate all the silicon under the antenna, but the resulting membrane structure is fragile and can be easily broken. In this thesis it has been shown that the etching of an array of small cavities results in a good gain improvement without compromising the structure robustness. Measurements show that not only the total amount of etched silicon is important, but the way how this silicon is eliminated is also relevant. The best etching patterns are the ones that eliminate more silicon near the feeding point and create horizontal trenches. With an etching depth of 450 μm, gain improvements up to 4 dB have been measured.

The antenna will be fabricated in a chip together with other electronic circuits that can degrade the RF link performance. This degradation has been analyzed with a particular example: the degradation introduced by a coil necessary for an inductive energy coupling. This is one of the worse cases because coil dimensions are similar to the antenna. Measurements show that the gain degradation depends on the separation between both elements. When this separation is reduced to obtain smaller chips, the RF link degradation is greater, with measured degradations of near 7 dB when the separation between both elements is 320 μm.

Some sensors can be implanted inside the human body. The analysis of this scenario is complex due to the presence of the body tissues. In simulations, complex numerical models have to be used to take into account the body effects. In measurements phantoms filled with body tissue simulating liquids are needed to emulate the real operating scenario of the antenna.

Chapter 4 studies the body effects when the radiating element is implanted. The main effects of having the antenna inside the body are a reduction of the antenna dimensions and a degradation of the antenna gain due to the human tissues high permittivity and conductivity values. The body effects in different ISM bands have been analyzed, showing that there exists a trade-off between antenna dimensions and system performance. When frequency is increased, antennas are smaller but the gain is lower, with differences near 20 dB when the antenna depth is 10 mm. Simulations also show that the antenna behavior

can change depending on the tissue layer where it is implanted. If it is in the fat layer, antenna gain is better, but it has to be larger to keep the same resonance frequency.

The main limitations of the one layer model have been analyzed. From the point of view of propagation losses the one layer model can be accepted as a conservative approach because it tends to overestimate the losses. However this model does not take into account the antenna detuning when it is moved from one tissue layer to another. Therefore, if the antenna design is carried out with the one layer model and finally it is implanted near or inside the fat layer, detuning problems can appear. For this reason multilayer models are recommended for the design of implantable antennas.

The effects of an insulation layer, necessary for biocompatibility issues, have also been analyzed. The insulation layer improves the antenna gain and antenna dimensions are less sensitive to the layer where it is implanted. The main drawback is that the antenna is larger. The propagation losses study of this chapter is based on simulated results, however the validity of FEKO simulations has been demonstrated with measurements of a patch antenna immersed in a body tissue simulating liquid.

Finally, it is important to point out that the results of this thesis have permitted the author to publish 5 journal papers and 7 conference papers (1 best paper award).

5.2 Future work

In this thesis the design of multiband antennas and silicon integrated antennas for a sensor network have been analyzed. The thesis is finished, but research never ends. For this reason, a list of topics that could be studied in future is given here:

- Use of other fractal shapes with self-similarity properties for the design of multiband antennas.
- Use of the resonator loading techniques with other kinds of antenna, like monopoles with small ground planes or PIFA antennas.
- Combine the use of fractal shapes with resonator loaded structures to improve multiband behavior.
- Study of multiband antennas performance degradation when they are in the vicinity of human tissues.
- Use of other fabrication processes (not necessarily compatible with standard CMOS

technologies) to improve antenna performance. It is important to evaluate the trade-off between performance and costs.

- Implementation of a measurement setup to carry out measurements of silicon integrated antennas outside the probe station.
- Use of an external antenna coupled to the chip (e.g. with an inductive coupling) to improve the radiation parameters without the need for physical connections between the chip and the external antenna.
- Study of silicon integrated antennas performance degradation when they are implanted inside the human body.
- Design of implantable antennas optimized for the 2.45 GHz and 5.8 GHz ISM bands.

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Fabrication and application of flexible, multimodal light-emitting devices for wireless optogenetics

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The rise of optogenetics provides unique opportunities to advance materials and biomedical engineering, as well as fundamental understanding in neuroscience. This protocol describes the fabrication of optoelectronic devices for studying intact neural systems. Unlike optogenetic approaches that rely on rigid fiber optics tethered to external light sources, these novel devices carry wirelessly powered microscale, inorganic light-emitting diodes (μ-ILEDs) and multimodal sensors inside the brain. We describe the technical procedures for construction of these devices, their corresponding radiofrequency power scavengers and their implementation *in vivo* for experimental application. In total, the timeline of the procedure, including device fabrication, implantation and preparation to begin *in vivo* experimentation, can be completed in ~3–8 weeks. Implementation of these devices allows for chronic (tested for up to 6 months) wireless optogenetic manipulation of neural circuitry in animals navigating complex natural or home-cage environments, interacting socially, and experiencing other freely moving behaviors.

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INTRODUCTION

Optogenetics is a relatively new field of neuroscience that gives researchers the ability to control cellular signaling and neural activity in a cell type–selective manner. *In vivo* applications of optogenetics have rapidly aided in the understanding of neural circuit function in behavioral models[1–10]. Despite the success of these studies, tethered fiber-optic approaches have restricted opportunities for the study of more complex, ethologically relevant behavioral paradigms such as enclosed home-cage behavior, spontaneous pain, wheel running and freely moving social interactions. Here we present a protocol for the fabrication of flexible devices that carry wirelessly powered microscale μ-ILEDs and multimodal sensors to study neural circuitry in awake, freely moving animals. The devices described in this protocol are robust, self-contained, multifunctional and they are capable of wireless operation with conventional electronics and power supplies. These fully electronic systems eliminate the need for high-powered light sources, fiber coupling fixtures and optomechanical hardware for *in vivo* optogenetic experiments. The following protocol is based on technology and approaches developed jointly in our two laboratories[11–16].

Development of the protocol and comparison with traditional light sources

This protocol is the result of advances in material science that have led to the development of flexible electronics, biodissolvable adhesives, microscale sensors and high-efficiency μ-ILEDs[12–14,17,18]. Although other groups have successfully implemented wireless schemes and LEDs for optogenetics[19–22], the protocol described here provides a completely customizable approach for combining various materials engineering approaches to design and implement devices that can be optimized for an individual laboratory's experimental needs.

The current standard in neuroscience for light delivery into the depth of the brain is to use chronically implanted fiber optics[23]; this approach offers substantial advantages over acute delivery of fibers via metal cannulae[1,11]. These chronic implants, however, have their own limitations. Principally, light from fiber-optic implants can only escape from the tip of the implant to illuminate ventral brain structures. Although adaptations to this ventral light delivery are possible[10,24,25] (available commercially at http://www.doriclenses.com), the range of customizability can be limited and often restricts the user to only delivering light without the capability of observing physiology. Furthermore, μ-ILEDs create opportunities to restrict or expand spatial targeting by selecting from a range of sizes (625–10,000 μm^2), by altering the number and arrangement of μ-ILEDs, and by using reflective materials to direct light. This protocol provides a basis from which any combination of μ-ILEDs and sensors can direct light within the brain and measure physiological function without the restriction of enforced light trajectory.

Various strategies have been used for delivering multiple wavelengths of light into the same animal[24–26]. These approaches require establishing an extensive network of optics and tethered optical equipment external to the behaving animal. Depending on the laboratory behavioral space, such setups can restrict experimental possibilities, and they can require advanced experience

with optics to maintain optimal conditions. This protocol may require access to external facilities for some laboratories, but the end result is a device that can be operated with basic laboratory equipment that is already likely to be present in most neuroscience laboratories. Furthermore, the rescue and recycling of these devices for reuse is relatively easy, meaning that these devices can be used for several rounds of behavioral experimentation with different sets of animals. For light-evoked activation of channelrhodopsin-2 [ChR2 (H134)], optically sensitive seven-transmembrane domain receptors (i.e., OPTO-α_1, OPTO-β_2), and other blue light–sensitive optogenetic constructs, GaN μ-ILEDs are appropriate[2,11,27]. A device that uses these 450-nm-emitting μ-ILEDs is the focus of this protocol. However, it is important to note that μ-ILEDs emitting at other relevant wavelengths are also possible to fabricate for use in other contexts[11,28]. The combination of μ-ILEDs of different wavelengths provides the user with access to activation spectra of multiple optogenetic constructs with a single implanted, electronic device. Furthermore, the electronic nature of these devices ensures that they can be operated wirelessly. Wireless optogenetic manipulation of neural circuitry has been achieved by other means[20,21], but these approaches can restrict behaviors accessible to study because the animal is required to remain in a fixed environment. By using radiofrequency (RF) power scavenging, the devices and approaches contained in this protocol free the user of constraints on behavioral assays, thereby allowing for experimental testing in any space.

Applications of the protocol

Although the focus of this protocol is on the creation of devices engineered to deliver μ-ILEDs into the brain for optogenetic applications, this same protocol can be used to fabricate devices to measure electrophysiological, thermal and other properties of intact brain tissue (**Box 1**). The flexible nature of these devices provides the potential to extend their application into other intact tissues such as the peripheral nervous system and the circulatory system/cardiac tissue of larger organisms.

Experimental design

Subjects. This protocol and these devices have been optimized for use in adult (25–35 g) male C57BL/6J mice and mutant mice backcrossed to the C57BL/6J mouse strain. However, as the optogenetic toolbox expands to other mammals[29–34], these devices will probably have broader utility in other animal models. Specifically, larger organisms such as rats and nonhuman primates will tolerate the ~700-mg wireless antenna with greater ease than mice. Unlike animals with polished fiber-optic implants, animals with chronically implanted μ-ILEDs can be housed with other animals, as the metal pin connectors cannot be damaged by cage mates. Before beginning this protocol, all the procedures described herein should be approved by the Animal Care and Use Committee of the investigating institution and conform to US National Institutes of Health (or other relevant governmental guidelines) guidelines regarding animal research. For this protocol, all procedures were approved by the Animal Care and Use Committee of Washington University and conformed to US National Institutes of Health guidelines.

Controls. As noted elsewhere[2,11,23], the proper controls for *in vivo* optogenetics are mice that express genetically encoded fluorescent reporters in the absence of an opsin. These control mice will account for any disturbance from viral injection, device implantation, heating confounds from light delivery and fluorophore fluorescence within the brain. Another important confound to consider when conducting light-evoked single-unit electrophysiological experiments is the potential for generating neural activity downstream of retinal stimulation. These devices prevent the external light escape that is common with fiber-optic implants, but activity from the visual system has the potential to activate the retina and retinal ganglion cells from within the brain, depending on the brain region[35]. Recordings performed with external light stimuli can control for these effects. In addition, properly counter-balanced within-subject experimental designs are possible with these devices by withholding power to the devices during the behavioral testing period.

Limitations. Although these devices offer numerous advantages over traditional fiber-optic implants, the technology is not intended to replace protocols that work well to manipulate neural circuitry *in vivo*[1,23]. The greatest limitation associated with this procedure is the accessibility to the facilities and the technical skill required to fabricate these devices and their scavenging antennas. That being said, subsequent recovery and refabrication of the devices is straightforward and accessible to any laboratory. Once the protocol is followed, there are a few considerations when designing behavioral experiments. First, depending on the panel antenna used, the signal can be polarized. In current designs, both the panel antenna and the antenna on the scavenger are directional with linear polarization. A maximum efficiency of power transmission and reception is achieved when the polarizations of both antennas are aligned. The power transmitted by the panel antenna can typically be adjusted to accommodate mismatches that can occur in most practical situations. Such issues can be avoided entirely by the use of a transmission antenna with circular polarization or multiple-panel antennas. Care should also be taken to ensure that the animal's surroundings do not interfere with the RF signal by powering a free-standing device in the behavioral context[11]. Wireless control of these devices is compatible with a wide range of behavioral assays. In our hands, the devices have performed well in tests of anxiety-like behavior (open field test, elevated zero maze), reward-related behavior (conditioned place preference, operant behaviors and self-stimulation), social behaviors (social defeat stress and social aversion), pain behaviors (Hargreaves test) and home-cage behaviors[11]. In our experience, commonly used materials for behavioral apparatuses (wood, polyvinyl chloride, poly(methyl methacrylate) and the metal from cage lids) do not interfere with the RF signal, but they can reduce the overall power if they are positioned between the panel antenna and the headstage scavenger (see TROUBLESHOOTING).

The current protocol only provides wireless access to power the μ-ILEDs and not to receive information from the sensors. This backward data-stream restriction is largely based on the weight that an animal can endure on the headstage. For a mouse, electrophysiological sensors, temperature sensors and photosensors currently require a wired connection. For larger mammals, the devices can be integrated with existing telemetry setups (commercially available at http://www.plexon.com), provided that they do not interfere with the RF signal[36,37].

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Box 1 | Combining multifunctional sensors and optoelectronics ● TIMING 1–4 d

One hallmark of these devices is their ability to both deliver light into the brain and to record information from the brain. Depending on the experiment, users may wish to outfit the devices with a variety of multiple and optional modalities. Each different modality can be combined with any other modality; devices can have between one and four different modalities incorporated. However, if the electrophysiological sensor is incorporated it should be the outermost layer to have direct exposure to the tissue.

1. The following steps provide information on adding functional layers (temperature sensors, option A, see also ; μ-IPDs, option B; and μ-electrodes for electrophysiological sensors, option C) to the devices. If these optional modalities are desired, **Box 1** should be completed before Step 16 of the PROCEDURE.

(A) Fabrication of temperature sensors

(i) Generate PR patterns on a needle-shaped piece of PET (same substrates used in Step 12) and deposit platinum (Pt, 10 nm) by sputtering. Remove PR by acetone and generate Cr/Au (15/300 nm) metal lines that connect to the Pt resistor.

(ii) Calibrate the temperature sensor. Dip the temperature sensor into DI water and measure reference resistance using a digital multimeter. With precisely controlled temperature of the DI water, measure the changes in resistance. As the resistance change (ΔR) of the Pt resistor depends linearly on the temperature change (ΔT), the formula for estimated temperature is $\Delta T \sim k \times \Delta R$, where k is constant.

(B) Fabrication of μ-IPDs

▲ CRITICAL (2nd layer shown in **Fig. 3**) Additional details of this method are shown in a previous publication[50].

(i) Deposit SiO_2 by PE-CVD on a silicon-on-insulator (SOI) wafer, with a top silicon layer thickness of ~1 μm.

(ii) Generate PR patterns to pattern the SiO_2 with HF.

(iii) Remove the PR and clean the wafer surface by RCA cleaning. First perform SC (standard cleaning)-1 with 1:1:5 solution of ammonium hydroxide (NH_4OH):hydrogen peroxide (H_2O_2):DI water at 80 °C for 15 min. Second, remove the SiO_2 layer using 1:50 solution of HF:DI water. Finally, perform SC-2 with a 1:1:6 solution of hydrochloric acid (HCl):H_2O_2:DI water at 80 °C for 15 min (RCA cleaning: http://inside.mines.edu/fs_home/cwolden/chen435/clean.htm).

(iv) Carry out solid-state doping of the silicon to form a p-type region.

(v) Repeat steps i–iv for n-type doping.

(vi) Generate PR hole patterns on the entire wafer surface and etch the silicon layer, to define the lateral dimensions of the photodetectors.

(vii) Etch the buried oxide (BOx) layer using HF.

(viii) Retrieve an ~1-μm-thick Si membrane photodiode using a PDMS slab and release it onto the needle-shaped PET substrate (described in Step12).

(ix) Form metal (Cr/Au, 15/300 nm) lines for interconnection.

(B) Fabrication of μ-electrode for electrophysiological sensor

(i) Generate PR (AZ 2030) lift-off patterns (1st layer shown in **Fig. 3**).

(ii) Deposit Pt (30 nm), a needle-shaped piece of PET (described in Step 12), by sputtering. Generate a pattern of SU-8 2 (2 μm thick) with a 20 × 20 μm^2 square opening.

▲ CRITICAL STEP An additional oxygen de-scum to the remove residual PR layer is recommended. The impedance of the Pt μ-electrode should be ~1.0 MΩ at 1 kHz.

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Finally, independent control of individual μ-ILEDs is also possible with these devices, but this protocol has not been optimized for these conditions in a wireless control mode. If independent control is a desired feature, it is recommended that smaller connections featuring a number of channels be used over the pin connectors that we present here (many options are commercially available at http://www.omnetics.com).

Upon completion of the protocol presented here, users can expect to have devices with four μ-ILEDs that can be wirelessly powered with suitable light output from 2 m away. In our experience, more μ-ILEDs can be added to a single device with relative ease (the maximum we have tested is 25 (5 × 5) μ-ILEDs from a single power scavenger). Furthermore, the RF power-scavenging approach is appropriately suited for powering multiple μ-ILED devices (presumably in multiple animals) by using a different power-scavenging headstage antenna for each device. The upper limit on the number of simultaneously powered devices is primarily dependent on the spatial constraints of the experimental space and apparatus to achieve equal powering to all devices.

Notes on the materials and equipment used in this protocol. Accurate injection of these devices into brain tissue requires a stereotaxic adaptor for standard cannula holders. The basic principle of this adaptor is to allow the μ-needle to be mounted in line with the existing stereotaxic system. The dimensions of such an adaptor will vary by stereotaxic alignment system make and model, but most university machine shops should be able to fashion such an adaptor. The adaptor presented in this protocol was specifically designed for use with the KOPF single cannula holder (model 1966). It is also possible that other commercially available electrode holders may be capable of accurately targeting the devices (e.g., KOPF model 1768). More detailed information on the machining and dimensions of the adaptor for the KOPF model 1966 can be found in **Supplementary Figure 1**.

To deliver pulse trains of light, the devices can be controlled using a traditional function generator to drive amplitude modulation of the wireless powering equipment. Amplitude modulation is an internal function available on most RF generators. This function allows an internally or externally supplied modulating

signal to control the amplitude of the output RF signal. The internal modulating signal is usually a sinusoidal waveform with a much lower frequency compared with that of the RF output. For the positive region of the sinusoidal signal, the RF generator can output an RF signal whose amplitude is modulated by the low-frequency sinusoidal signal, and it can change from 0 to the maximum set power and then back to 0. For the negative region of the modulating signal, the output RF signal maintains at 0. An external transistor-transistor logic (TTL)-modulating signal is preferable in terms of modulating the amplitude of the output RF signal. Here, the high state of the TTL causes the RF generator to output a constant set power, whereas in the low state of the TTL the RF generator outputs 0 power. Thus, the TTL-modulating method is more suitable for generating constant light intensity during the high state of the TTL. The pulse width and frequency of photostimulation should be determined and based on physiologically relevant conditions[7,10,38–40].

Postmortem rescue of the devices for reuse is a delicate but straightforward process. The dental cement presented here (Lang Dental) allows for such rescue. Other cements and bonding agents can inhibit the process and destroy internal components. It may be necessary to use a stronger bonding agent in some scenarios (see TROUBLESHOOTING). In such cases, the reusability might be compromised.

MATERIALS

REAGENTS

Preparation of releasable polymer template

- UV-curable epoxy (SU-8 2 and 100 photoresist (PR); Microchem)
  ! CAUTION Partially cured or uncured epoxy needle can induce excess chemical contamination in the brain.
- SU-8 developer (Microchem)
- Isopropyl alcohol **! CAUTION** Isopropyl alcohol is flammable.
- Acetone **! CAUTION** Acetone is flammable.
- Silk adhesive (details are shown in other papers[13,17,18]; see Reagent Setup)
- Benzocyclobutene (BCB, DS-4022 35; Dow Corning)
- BCB developer (advanced developer, DS2100, Dow Corning)
- Water-soluble tape (3M)
- PRs (AZ 1518, AZ 2070; Capital Scientific)
- AZ 300 MIF developer (Capital Scientific)
- Hydrofluoric acid (HF) **! CAUTION** HF is extremely corrosive. Wear gloves and use eye protection when using HF.
- Chrome (Cr) etchant, gold (Au) etchant, palladium (Pd) etchant (Transene)
- Polydimethylsiloxane (PDMS; Sylgard 184, Dow Corning)
- Polyethylene terephthalate (PET; polyester film, 2.5 and 6 μm thick, Mylar film, Chemplex)

Preparation of μ-ILEDs

- Gallium nitride (GaN) LED epitaxial materials grown on sapphire substrates (Cermet)

Injection of virus and μ-ILEDs into targeted brain structure

- Isoflurane, USP (Isothesia, Butler Schein, cat. no. 029405) **! CAUTION** Before beginning this protocol, the Animal Care and Use Committee of the investigating institution should approve all procedures and conform to US National Institutes of Health (or other relevant governmental guidelines) guidelines regarding animal research **! CAUTION** Ensure that proper ventilation and gas scavenging methods are in place to prevent potential inhalation of excess isolflurane.
- Betadine solution (Purdue Products, cat. no. 67618015017)
- Ethanol (Sigma-Aldrich, cat. no. 362808) **! CAUTION** Ethanol is flammable.
- Hydrogen peroxide, 3% (wt/vol) USP (Select Medical Products, cat. no. 117)
- Lidocaine ointment USP, 5% (wt/vol) (Fougera)
- Ophthalmic ointment (Altalube Ophthalmic Product)
- Adhesive luting cement (C&B-Metabond, Parkell, cat. no. S380) **! CAUTION** Methyl methacrylate monomer, stabilized is a flammable liquid. It may also cause skin irritation; avoid contact with skin, eyes and clothing. Use it with adequate ventilation.
- Light cured bonding adhesive (VLC one-etep prime and bond adhesive, cat. no. 305-006-030) **! CAUTION** Methyl methacrylate monomer, stabilized is a flammable liquid. It may also cause skin irritation; avoid contact with skin, eyes and clothing. Use it with adequate ventilation.
- Dental cement (Jet denture repair, Lang Dental, cat. no. 1223, see Reagent Setup) **! CAUTION** Methyl methacrylate monomer, stabilized is a flammable liquid. It may also cause skin irritation; avoid contact with skin, eyes and clothing. Use it with adequate ventilation.
- Sodium chloride injection, 0.9% (wt/vol), USP, 9 mg ml^{-1} NaCl (Hospira, cat. no. RL-0497(9/04))
- Enrofloxacin (Baytril, Bayer, cat. no. R30901)
- Antibiotic ointment (Neosporin; Johnson & Johnson, cat. no. 174-73087Q)
- Viruses of interest (adeno-associated virus and lentiviruses are available from the Washington University of St. Louis Hope Center Viral Core, https://hopecenter.wustl.edu/?page_id=99 and/or the University of North Carolina Viral Vector Core, http://genetherapy.unc.edu/services.htm). Herpes simplex viruses are available from the Massachusetts Institute of Technology (MIT) Viral Core, http://mcgovern.mit.edu/technology/viral-core-facility)
  ! CAUTION Follow the appropriate safety precautions pertaining to the particular virus in use. It may be necessary to obtain a higher Biosafety Level certification before use.
- Artificial cerebral spinal fluid (ACSF; Tocris, cat. no. 3525 or custom, see Reagent Setup)[1,41]
- Animal(s) to be injected with the μ-ILED device. The procedure describes how to use a μ-ILED device with a mouse, as previously described[11]
  ! CAUTION Before beginning this protocol, the Animal Care and Use Committee of the investigating institution should approve all procedures and conform to US National Institutes of Health guidelines (or other relevant governmental guidelines) regarding animal research.

Behavioral procedures using RF power scavenging

- Thermal grease (Wakefield Solutions, cat. no. 120-2)

EQUIPMENT

Preparation for μ-ILEDs and multifunctional sensors

- Sputter (AJA international, ATC 200) metal deposition for Au (gold), Ni (nickel), Pt (platinum), etc.
- Rapid thermal annealing (RTA) **! CAUTION** High temperature, semi-transparent L-shaped current spreading layer on p-side on GaN LED should be used in thin metal layers (15/15 nm Ni/Au) followed by annealing at 500 °C.
- Mask aligner (Karl Suss, MJB)
- Inductively coupled plasma etcher (ICP; PlasmaTherm SLR-700, etching for GaN)
- Reactive ion etcher (RIE, PlasmaTherm 790, etching for SiNx)
- Plasma enhancement-chemical vapor deposition (PE-CVD; STS-mixed frequency nitride deposition system)
- Wafer bonder (Electronic Visions, EV501)
- Laser lift off (LLO; Krypton Fluoride (KrF), 0.9 J cm^{-2}, 248 nm wavelength (IPG Photonics (http://www.ipgphotonics.com/microprocessing.htm)) or Yttrium aluminum garnet (YAG):Nd laser (0.3 J cm^{-2}, 266 nm single pulse with 5 ns exposure, Sandia National Lab) **! CAUTION** Avoid eye contact with the laser; always wear eye protection.
- Digital multimeter (Fluke 115 or other commercially available multimeter)
- Printed circuit board (PCB) for headstage connection (General Circuits)
- Male pins for headstage PCB, 2.54-mm spacing male pins (Sunlight Inc., cat. no. P2540-H254-S180)

Measurement of μ-ILED

- Ocean Optics (Ocean Optics, HR4000; for measuring wavelength spectrum, light output power)
- Probe station (Agilent 4155; current-voltage (I-V) characteristics)
- Pulse generator (Global specialties, 4001)

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- Oscilloscope (Agilent, DSOX2004A, 70 MHz)
- IR camera (QFI InfraScope II)

Injection of virus and μ-ILEDs into targeted brain structure

- Stereotaxic alignment system (KOPF model 1900) ! CAUTION Before beginning this protocol, the Animal Care and Use Committee of the investigating institution should approve all procedures and conform to US National Institutes of Health guidelines (or other relevant governmental guidelines) regarding animal research.
- Stereotaxic alignment indicator (KOPF model 1905)
- Centering microscope, 40× (KOPF 1915)
- Stereotaxic drill (KOPF model 1911) and no. 66 drill bit (KOPF, cat. no. 8669)
- Stereotaxic single cannula holder (KOPF model 1966)
- Stereotaxic adaptor for cannula holder (see **Supplementary Fig. 1**, made in your institutional machine shop or other)
- Anchoring screws (CMA Microdialysis, cat. no. 7431021)
- Needles (Becton-Dickinson, cat. no. 305111)
- Microinjection syringes (Hamilton, cat. no. 88011)
- Infusion pump and controller (UltraMicroPump III, World Precision Intstruments, cat. nos. UMP3 and UNC4)
- Forceps (Miltex, cat. no. 6-100)
- Surgical scissors (Miltex, cat. no. 18-1430)
- Hemostats (Miltex, cat. no. MH7-26)
- Microspatula (Chemglass, cat. no. CG-1982-12)
- Electric clippers (Wahl, cat. no. 8064-900)

Behavioral procedures using RF power scavenging

- RF generator (Agilent, N5181 MXF)
- Function generator with standard TTL (AMPI Master-9 or other)
- RF amplifier (Empower RF systems, 1100/BBM2E4AJP)
- RF antenna (ARC wireless solutions, ARC-PA0913B01, ARC 902–928 MHz, 12.5 dBi, flat panel)
- DC power supply (Mastech, cat. no. HY5005E-2)
- Relevant behavioral assay apparatus (university machine shop, MED Associates, Harvard Apparatus or other)
- Wireless headstage antenna (see PROCEDURE)
- PCB for wireless headstage antenna (General Circuits)
- Ceramic antenna (W3012; Pulse Electronics)
- Schottky diode (Digi-Key, cat. no. MMDL301T1G)
- Power meter (Bird Electronic, Thruline 43)
- Two SMA-male to N-male precision cables (Pasternack, cat. no. PE304-120)
- Fan-cooled heat sink (Fischer Elektronik, cat. no. LA 17/200 24V)

Behavioral procedures using a wired connection

- Animal(s) with injected μ-ILED device ! CAUTION Before beginning this protocol, the Animal Care and Use Committee of the investigating institution should approve all procedures and conform to US National Institutes of Health guidelines (or other relevant governmental guidelines) regarding animal research.
- Relevant behavioral assay apparatus (university machine shop, MED Associates, Harvard Apparatus or other)
- Function generator for standard TTL (AMPI Master-9 or other)
- BNC cables (Cables to Go, various lengths and cat. nos.)
- Electrical rotary joint (Moog, cat. no. SRA-73683 (ref. 42))
- BNC-to-banana plug adaptor (Fluke cat. no. BP881 or other)
- Wire, 30 gauge (Artistic Wire) and female connector (single modality devices: TE Connectivity, cat. no. AMP 3-640441-2; multimodal devices or independent μ-ILED control: TE Connectivity, cat. no. AMP 87631-4)

REAGENT SETUP

Silk adhesive Boil cocoons of the *Bombyx mori* silkworm for 30 min in a solution of 0.02 M Na_2CO_3 to remove the sericin protein. Rinse the extracted fibroin with dH_2O and dry it in ambient air for 12 h. After drying, dissolve the fibroin in a 9.3 M LiBr solution at 60 °C for 4 h, yielding a 20 wt% aqueous solution, and subsequently dialyze the solution against dH_2O by using a dialysis cassette at room temperature (20–25 °C) for 48 h until the solution reaches a concentration of 8 wt%. Silk adhesive can be stored at room temperature for up to 6 months.

Dental cement The dental cement can be prepared as described elsewhere[23] by combining the Jet denture repair powder with the Jet liquid (methyl methacrylate). A starting point is 350 l of methyl methacrylate monomer added to 225 mg of powder. In some cases, the viscosity must be adjusted (by increasing or decreasing the amount of methyl methacrylate in the mixture) to properly secure the flexible device in place and/or mount the PCB in the headcap. Dental cement can be stored at room temperature for up to a year. ▲ CRITICAL Adjusting the viscosity alters the working time of the cement.

Virus Each type of virus has different handling instructions, but all must be stored on ice until immediately before injection. In some cases, dilution of stock titers requires obtaining more of the reagent in which the viral preparation was initially concentrated.

ACSF If not purchased, the solution can be prepared according to recipes described elsewhere[1,41]. It can be stored as a stock solution at 4 °C for up to 1 month and at room temperature as a working solution for 1 d.

EQUIPMENT SETUP

Preparation of amplifier with proper heat sink The RF amplifier requires an additional cooling system during operation. Thermal grease applied on top of the heat sink facilitates thermal contact with the amplifier, which is affixed to the sink using screws. The DC power supply operates both the amplifier and the fan attached to the heat sink by supplying power with voltages of 24 V and 10 V, respectively.

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PROCEDURE

Preparation of μ-ILEDs ● TIMING 7 d

1| Clean the wafer surface of GaN LED stacks (450-nm emission wavelength; p-type GaN/multi-quantum well (MQW)/spacer/n-type GaN/undoped GaN) grown on sapphire substrate with 5 wt% diluted HCl for 5 min.

▲ CRITICAL STEP Proper cleaning enables reduced turn-on voltages and improved efficiencies, both of which minimize the production of heat by the operating μ-ILEDs.

2| Deposit metals (Ni/Au, 15/15 nm) immediately after cleaning.

▲ CRITICAL STEP The deposition should be carried out at degree vacuum levels of ~10^{-6} torr or less.

3| Form L-shape patterned metals (Ni/Au) onto the p-type GaN semiconductor layer by photolithography and wet etching with Cr and Ni etchants; follow by removal of the PR with acetone (**Fig. 1a**).

4| Anneal the wafer for 10 min at 500 °C in 20% oxygen and 80% argon atmosphere to generate *p*-ohmic contact.

▲ CRITICAL STEP The annealed metal layers should be semitransparent after this step.

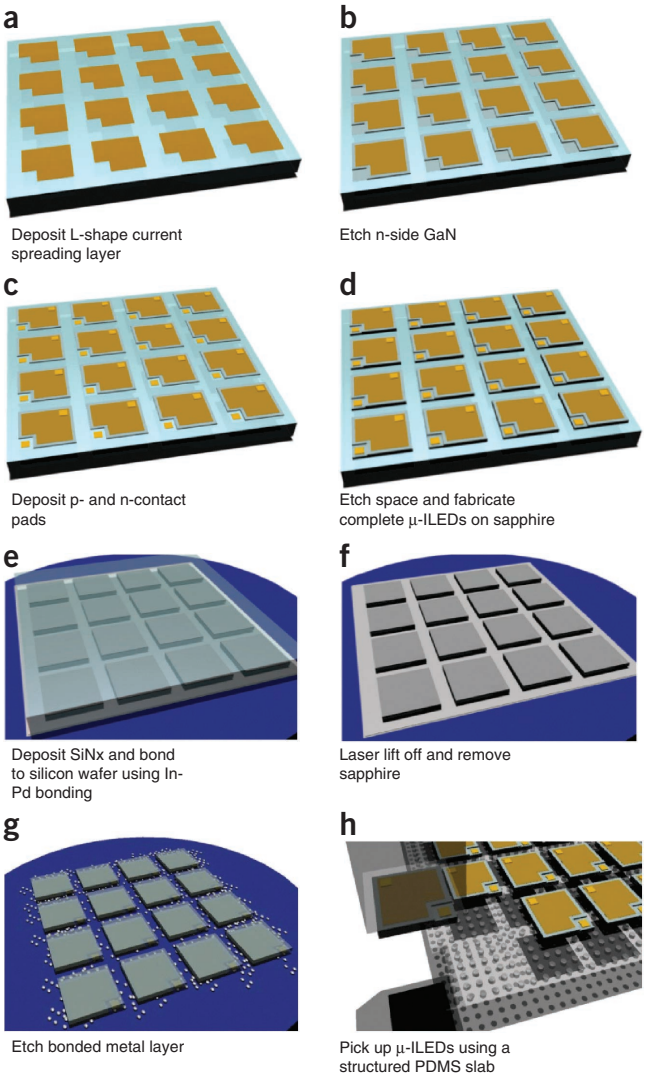

Figure 1 | Fabrication procedure for injectable μ-ILEDs. (**a**) An L-shaped current spreading layer (patterning of Ni/Au (15/15 nm) layers followed by 500 °C annealing) is formed on GaN blue lighting LED stacks grown on sapphire. (**b**) To expose n-type GaN semiconductors, etch rectangular shapes of p-type one, multi-quantum well (MQW) and spacer layers by reactive ion etching. (**c**) Generation of n- and p-contact pads (Cr/Au (15/300 nm) 25 × 25 μm^2 squares) on n-and p-type GaN sides. (**d**) Etch a 20-μm-width trench to define lateral dimensions of 100 × 100 m^2 or smaller. (**e**) Bonding μ-ILEDs on sapphire with a silicon wafer using an In-Pd metallic alloy bond. (**f**) Expose the laser on sapphire and mechanically remove the sapphire from μ-ILEDs. (**g**) Wet-etch the bonded metallic layer located on the wafer. (**h**) Transfer of all μ-ILEDs onto structured PDMS slabs and selectively pick up single μ-ILEDs by using a polydimethyl siloxane (PDMS) stamp. Panels **d**, **f**, **g** and **h** are reproduced from ref. 12 with permission from Wiley.

5| Generate 40 × 40 μm^2 square hole patterns in a negative-tone PR (AZ 2070 PR) and etch (chlorine gas–based RIE etching) the p-GaN layer, MQW, spacer to open n-side GaN (**Fig. 1b**).

6| Generate 25 × 25 μm^2 square hole patterns in PR for n- and p-type metal contact pads and deposit Cr/Au (10/300 nm) by using an electron beam evaporator. Then, remove PR (**Fig. 1c**).

7| Deposit a low-stress layer (200 nm) of SiNx on the substrate, by using PE-CVD. Next, pattern a negative-tone PR (AZ 2070) to serve as a mask for etching the SiNx and the GaN to define the lateral dimensions of the array of μ-ILEDs. Devices with dimensions of 100 × 100 μm^2 or less are formed with 20-μm spacings (**Fig. 1d**).

8| Deposit a layer of SiNx onto the μ-ILEDs to protect the devices and metal pads from further processes. Bond the LED wafer to a silicon substrate with an indium (In)-palladium (Pd) metallic alloy. This bonding uses Cr/Pd (15/150 nm) on the LED substrate and Cr/Pd/In (15/150/900 nm) on the silicon. The bonding occurs on contact with a pressure of 400 bar and a temperature of 220 °C for 2 h (**Fig. 1e**).

9| Use laser illumination through the sapphire to release the μ-ILEDs (**Fig. 1f**). Details of this LLO technique appear elsewhere[12,43].

10| Etch the unalloyed metal by immersion in 5 wt% HCl for 30 min (**Fig. 1g**).

11| Retrieve all μ-ILEDs onto the substrate of PDMS with patterns of surface relief, in the geometry of cylindrical pillars with 3 μm diameter and 1.4 μm height and 5 μm spacing. Remove residual metals, including In-Pd alloy and Cr, by Pd and Cr etchants (**Fig. 1h**).

12| Prepare a PDMS stamp with a single relief feature consisting of a post with 100 × 100 μm^2 lateral dimensions and 100 μm height. By using a mask aligner for photolithography, retrieve a single μ-ILED with the stamp and transfer it to a needle-shaped structure of PET adhered to a thin layer of PDMS on a transparent temporary substrate (**Fig. 2**).

13| Remove the passivation layer, SiNx (deposited in Step 8), with RIE and spin-coat a photosensitive BCB (6-μm-thick) onto the μ-ILED.

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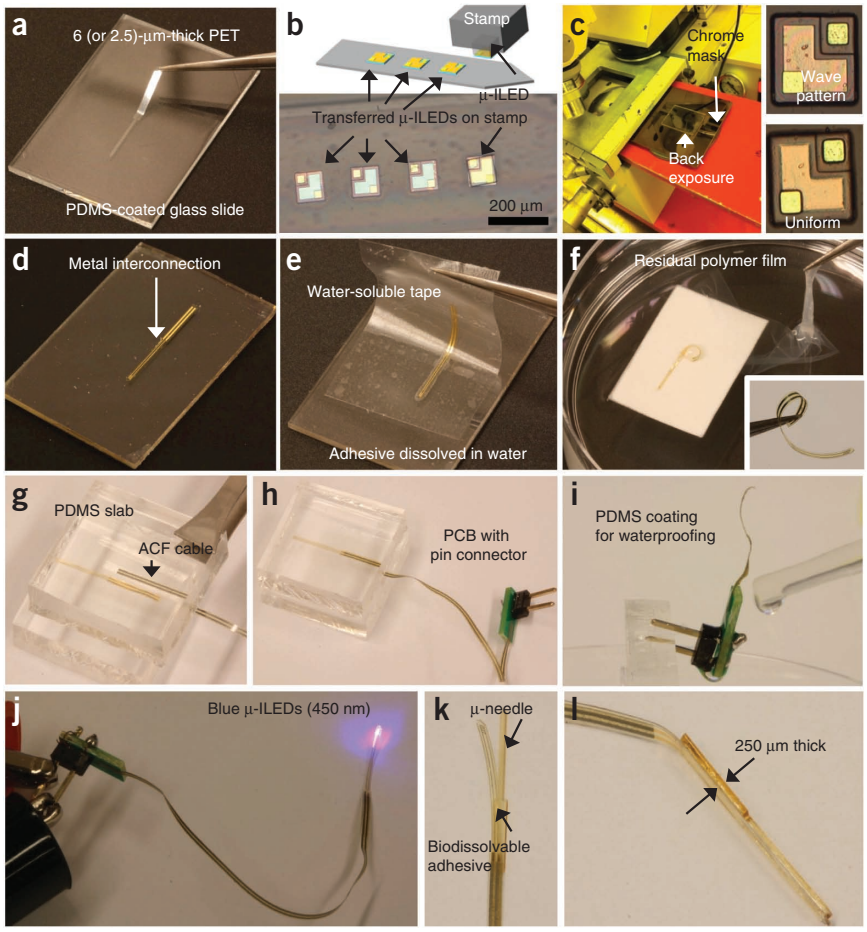

Figure 2 | Procedure for fabrication of injectable, multifunctional electronics. (**a**) Thin (~2.5-μm thick) needle-shaped PET is attached to a temporary PDMS-coated glass substrate. (**b**) Schematic and photograph demonstrating the transfer printing of four μ-ILEDs onto the tip of the PET using a PDMS stamp. (**c**) Passivation approach with photocurable BCB polymer. The back of the BCB-coated substrate is exposed to UV light. The wave pattern in the upper inset shows nonuniform coating of BCB. The lower inset shows successful uniform coating. (**d**) The metal interconnection (Cr/Au) is generated by sputtering, photolithography and metal etching to electrically connect the four μ-ILEDs. (**e**) The connected device is picked up with water-soluble tape. (**f**) The substrate is separated from the tape after the adhesive is dissolved in the water. The inset shows the μ-ILEDs on free-standing thin, flexible, needle-shaped PET. (**g**) The device is electrically connected to the ACF cable. The PDMS slabs on the top and bottom are compressed using high temperature (~150 °C) to bond the ACF cable. (**h**) The other side of the ACF cable is connected to the PCB with pin connector for wireless or wired powering schemes. (**i**) The ACF cable and PCB are coated with PDMS for waterproofing. (**j**) Blue (450 nm) μ-ILEDs are powered. The μ-ILEDs and electrical connection should be checked before injection. (**k**) The device is assembled with injection μ-needle using biodissolvable silk adhesive. (**l**) Image of a completed device ready for injection into brain tissue.

14| Pass UV light through the back of the substrate and develop the BCB with BCB developer (**Fig. 2c**) to obtain holes for n- and p-side contacts (**Fig. 2c**, lower inset). After further curing with BCB, anneal the samples at 250 °C on a hot plate for 3 h.

▲ **CRITICAL STEP** Make sure that BCB covers the μ-ILED surface when curing. Nonuniform coatings of BCB generate interference fringes that are readily visible (**Fig. 2c**, upper inset).

▲ **CRITICAL STEP** The annealing should be carried out in Ar atmosphere to avoid high-temperature damage to the BCB and polymeric layers.

? TROUBLESHOOTING

15| Deposit Cr/Au (15/300 nm) by sputtering, and form metal interconnects using positive-tone PR (AZ 15198), followed by metal etching (**Fig. 2d**).

Deterministic device assembly by transfer printing and formation of electrical interconnects ● TIMING 1 d

16| Remove μ-ILEDs fabricated on each temporary substrate, as shown in **Figure 2a**, with water-soluble tape (**Fig. 2e**). Remove the water-soluble tapes in deionized (DI) water (**Fig. 2f**) and place the devices on slabs of PDMS (**Fig. 2g**). If appropriate, follow instructions in **Box 1** to assemble a multifunctional device (**Fig. 3**).

▲ **CRITICAL STEP** Steps 16–20 describe the device assembly for μ-ILEDs alone. If you are assembling a multifunctional device, follow the instructions in **Box 1** before proceeding.

17| Connect each of the devices with anisotropic conductive film (ACF) cable to a PCB board (**Fig. 2h**).

▲ **CRITICAL STEP** Apply pressure and high temperature (~150 °C) during ACF cable bonding.

? TROUBLESHOOTING

18| Assemble all devices onto an injection μ-needle formed in SU-8 epoxy. The entire region of the injection device should be electrically insulated for *in vivo* use. Cover the flexible device with SU-8 polymer and the other parts with PDMS

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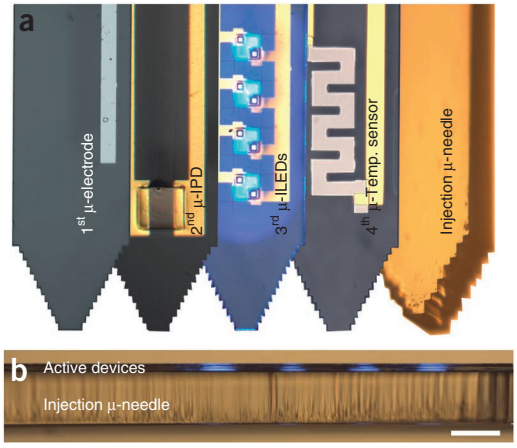

Figure 3 | Multifunctional sensors and optoelectronics. (**a**) Representative scheme for multifunctional, injectable electronics formed on an injectable needle. The devices include an electrophysiological sensor (μ-electrode; 1st layer), a silicon photodiode (μ-IPD; 2nd layer), four microscale μ-ILEDs (3rd layer), and a temperature sensor (μ-temperature (μ-temp.) sensor; 4th layer) based on a platinum resistor are formed on an injectable μ-needle fabricated from epoxy polymer. (**b**) Side view of such a device reveals the ultrathin nature of the active components of the device. Scale bar, 200 μm.

for waterproofing (**Fig. 2i**). Temporarily fix the device to the PDMS slab. Drop PDMS solution (Sylgard 180, Dow Corning, weight ratio = 10:1 (silicone elastomer precursor: curing agent), see http://www.dowcorning.com/applications/search/default.aspx?r=131en) onto the ACF cable and PCB, except for the pin connection area. After baking the PDMS in the oven at 70 °C for 2 h, the entire area except the pin adaptor is electrically passivated by PDMS.

19| To check functionality, power the device by connecting it to a power supply capable of delivering more than 1 mW (**Fig. 2j**). Use silk adhesive to bond the devices to the injection μ-needle (**Fig. 2k**,**l**).

Fabrication of wireless power harvester and preparation of wireless operation ● **TIMING 1 d**

20| Build either the robust, PCB-based (option A) or thin, flexible (option B) wireless power harvester.

▲ **CRITICAL STEP** Both options perform equally, but each has its advantages. The PCB-based option is more easily constructed and more durable, but it weighs more than the flexible version. We recommend using the PCB-based harvester for experiments that involve social interaction, whereas the flexible version is ideal for experiments involving only a single animal.

(A) Construction of the PCB-based wireless power harvester

(i) Build the wireless power harvester from two stacked PCB circuits (**Fig. 4**; general circuits). Each PCB includes top and bottom layers that are mostly covered by solid copper ground planes.

(ii) Mount the electrical components on the PCB boards through soldering. The first PCB circuit (labeled as 5 in **Fig. 4b**) contains a ceramic antenna operated at 915 MHz with a clearance space of 10.8 × 8.25 mm to the surrounding ground plane.

(iii) Connect a capacitor (8.2 pF in capacitance) between the feed line of the antenna and the ground plane to match the impedance of the antenna with the following circuit. A secondary PCB circuit (labeled as 6 in **Fig. 4b**) contains a voltage multiplier circuit constructed with six pairs of capacitors (47 nF in capacitance) and Schottky diodes (MMDL301T1G) in cascaded connection[11].

(iv) Mount the connector that is used for the μ-LEDs to the secondary PCB circuit. Join the PCB circuits and the connector with flexible copper wires (**Fig. 4c**), and then bond them together with resin epoxy (**Fig. 4d**). **Figure 4i** shows a schematic illustration of the power harvester. The circuit parts, which are labeled with numbers 5–7, correspond to the same numbers shown in **Figure 4b**,**e**.

(B) Construction of the thin, flexible wireless power harvester

(i) Fabricate the flexible PCB circuit on a 75-μm-thick Kapton film (**Fig. 4f**). Coat the film with a 6-μm-thick layer of copper by using electron beam evaporation.

(ii) Pattern the copper to form pads for the electrical components and interconnection. Spin-cast and pattern a 1-μm-thick polyimide film to expose the electrical contact pads, while passivating the interconnection (**Fig. 4f**).

(iii) Attach a ground plane made of copper tape (**Fig. 4g**) to the bottom of the Kapton film and connect it with the copper pattern on the top of the Kapton film through channels filled with solder. A schematic illustration of the flexible PCB circuit appears in **Figure 4j**.

21| Set up the wireless power transmission system. The system consists of an RF signal generator (N5181A, Agilent Technologies), an RF amplifier (1100/BBM2E4AJP, Empower RF Systems), a DC power supply (U8031A, Agilent Technologies), an antenna (ARC-PA0913B01, ARC Wireless) and a power meter (Thruline 43, Bird Electronic) (**Fig. 4a**). Connect the amplifier to the RF signal generator and an antenna by using coaxial cables (SMA-male to N-male precision cable with 160-series co-ax, Pasternack). The RF signal generator provides a 915-MHz RF signal with a power of −15 to −20 dBm. The signal is amplitude-modulated through the internal function of the generator to create a pulsed signal with a frequency of 10 Hz and

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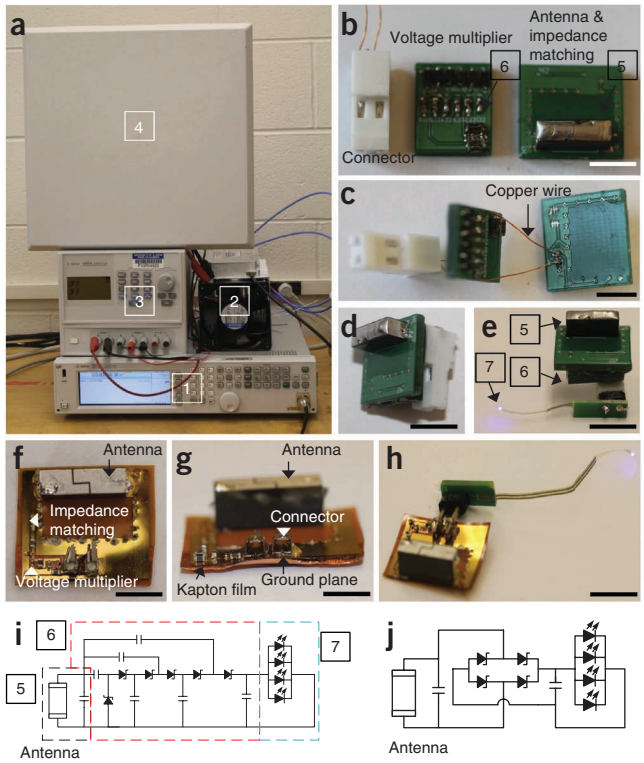

Figure 4 | Wireless operation and equipment. (**a**) An experimental setup for wireless power transmission. The setup contains an RF signal generator (1), an RF power amplifier (2), a DC power supply (3) and a panel antenna (4). Components for the wireless power harvester for μ-ILED powering with stacked PCB circuits (5: the circuit contains a ceramic antenna, and a capacitor is connected between the feed line of the antenna and the ground plane to match the impedance of the antenna with the next circuit. (**b**,**c**) A second circuit contains a voltage multiplier constructed with six pairs of capacitors and Schottky diodes in a cascaded connection(6)), before (**b**) and after (**c**) connecting with a copper wire. (**d**,**e**) A completed wireless power harvester alone (**d**) and with (**e**) connection to the μ-ILED device (7) for wireless operation. (**f**,**g**) Top (**f**) and side views (**g**) of a flexible wireless power harvester on Kapton film with similar components as the wireless harvester on PCB circuits. (**h**) A completed flexible wireless power harvester with connection to the μ-ILED device for wireless operation. Scale bars, 5 mm. (**i**) A schematic of the PCB-based power harvester. The numbered circuit components correspond to the same numbers shown in panels **b** and **e**. (**j**) A schematic of the wireless power harvester.

a carrier frequency of 915 MHz. Alternatively, an external TTL function generator can be connected to the RF signal generator to modulate the pulse width and frequency. This signal is directed to the RF amplifier, which amplifies the signal to 30–33 dBm.

22| Connect the power supply to the RF amplifier and the exhaust fan of the heat sink of the amplifier (see Equipment Setup) to provide DC voltages of 24 V and 10 V to the RF, respectively. Measure the intensity of the output power of the amplifier, and then connect the amplifier to the antenna to RF power to power the μ-LEDs at a 1~2 m distance (**Fig. 4e**,**h**).

Injection of the virus and μ-ILEDs into targeted brain structure ● TIMING 1 d, plus expression time

23| Mount the μ-ILED device in the custom-built cannula holder adaptor or another electrode or device holder (see Equipment Setup) (**Fig. 5a** and **Supplementary Fig. 1**). To do so, grasp the exposed region of the μ-needle (**Fig. 5b**).
▲ CRITICAL STEP The device must be placed along the midline of the adaptor to achieve proper spatial targeting.

24| Anesthetize the mouse in an isoflurane induction chamber using 4% (wt/vol) isoflurane and an oxygen flow rate of 1.5 liters per min.
! CAUTION Before beginning this protocol, all procedures should be approved by appropriate Animal Care and Use Committees, and they should conform to institutional and governmental regulations.
▲ CRITICAL STEP Ensure that the animal is sufficiently anesthetized before transferring it to the stereotaxic frame. The animal should have no response to a toe pinch, and its breathing rate should reduce to ~1 Hz.
? TROUBLESHOOTING

25| Transfer the animal to the stereotaxic frame, ensuring proper airflow of isoflurane to the nose cone (~2.5% (wt/vol) isoflurane; 1.5 liters per minute O_2).
▲ CRITICAL STEP Isoflurane levels should be monitored throughout the surgery to maintain sufficient anesthesia (no toe-pinch response) and breathing (1 Hz). For longer-duration surgery, isoflurane levels may be decreased to as low as 1% (wt/vol). Sterile, aseptic conditions should be used at all times to avoid infection.
▲ CRITICAL STEP Proper placement in the stereotaxic frame is required for accurate injection of virus and devices.
? TROUBLESHOOTING

26| Apply ophthalmic ointment to protect the animal's eyes during surgery.

27| Shave a 2 × 1 cm area on the animal's scalp (**Fig. 5c**).

28| Inject 0.1 ml of enrofloxacin into the hind limb (i.m.) and 0.3 ml of saline solution (0.9% (wt/vol), s.c.) to prevent infection and dehydration, respectively.

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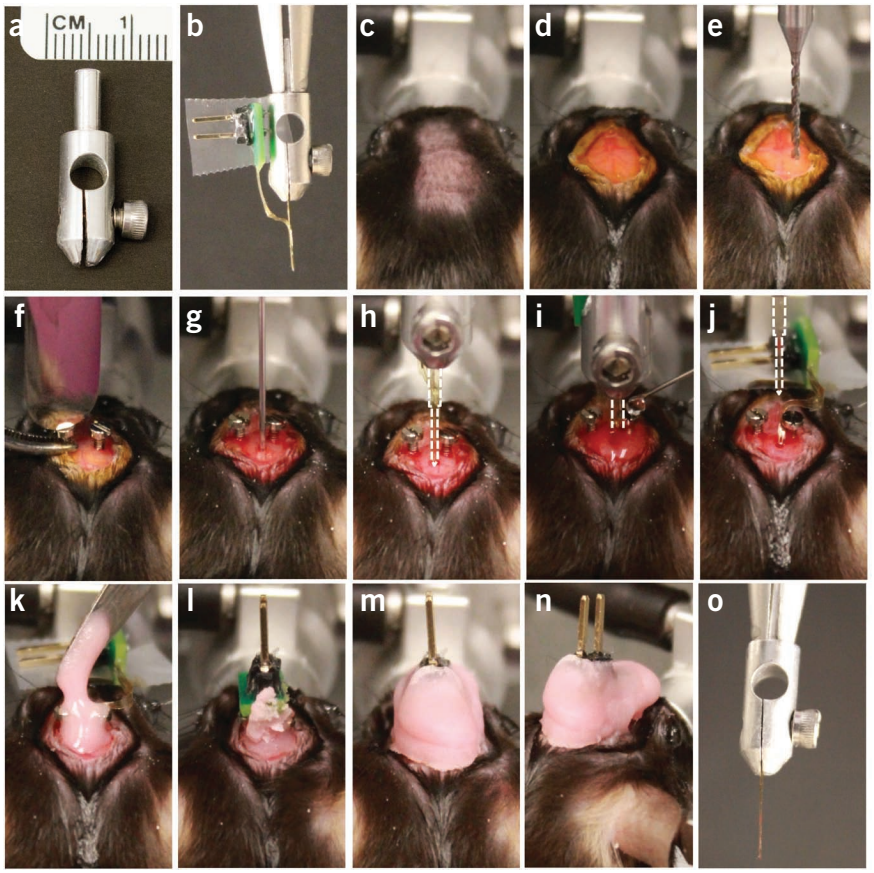

Figure 5 | Surgical procedure for injection of virus and μ-ILED devices into mouse brain. (**a**) Custom-built adapter for accurate stereotaxic placement of device (see Equipment Setup). (**b**) Mounted μ-ILED device, ready for injection into the animal. The exposed μ-needle is grasped with the adapter and a small piece of tape is used to secure the PCB during surgery. (**c**) A properly mounted mouse, with head shaved and eyes lubricated, is ready for surgery. (**d**) Betadine and ethanol are used to prevent infection and the scalp is open to expose the skull. (**e**) After leveling the skull, the drill is used to create pilot holes for the bone screws. (**f**) Forceps and a spatula or jewelry screwdriver are used to drive the screws into the skull. (**g**) The syringe needle is lowered to the desired coordinates to deliver the virus containing the optogenetic construct. (**h**) A μ-ILED device prepared to be driven into the brain using the same craniotomy as the viral injection. Dashed lines outline the shape of the device for clarity. (**i**) The μ-ILED device is lowered into the tissue and ACSF is applied to the skull surface to dissolve any external silk adhesive. (**j**) After a 15-min waiting period, the μ-needle is carefully retracted from the skull. (**k**) Dental cement is applied directly to the craniotomy site to secure the μ-ILED device in its targeted position. (**l**) The PCB connector is secured above the bone screws using a second layer of dental cement. (**m**,**n**) The PCB connector is completely encapsulated in dental cement, taking care to ensure that no bonds are made directly to the soft tissue. (**o**) The adapter is shown after surgery, containing only the μ-needle. All procedures were approved by the Animal Care and Use Committee of Washington University and conformed to US National Institutes of Health guidelines.

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29| By using a cotton swab, apply 70% (vol/vol) ethanol and Betadine to the shaved area.

30| Carefully grasp the tissue with the forceps and use the scissors to make an incision the length of the shaved area of the animal's scalp (**Fig. 5d**).
▲ **CRITICAL STEP** Ensure that all surgical instruments are sterilized before each animal surgery.

31| After the skin separates, use the forceps and scissors to clean and remove any remaining periosteum on the skull surface.

32| Identify the skull sutures, lambda and bregma. Carefully level the skull according to the manufacturer's recommended steps for your stereotaxic frame. For the KOPF model 1900, use the stereotaxic alignment indicator. It is crucial that both lambda and bregma be level, and you must achieve a level plane laterally across the skull.
? TROUBLESHOOTING

33| Once the skull is level, move the drill to the coordinates you have selected for anchor screw placement. Carefully drill a hole with a diameter just wide enough for the screws to catch and not so deep as to penetrate the skull (**Fig. 5e**).
▲ **CRITICAL STEP** Proper anchoring is crucial to ensure that the headcap remains affixed to the skull for the duration of the behavioral experiments. Generally, anchor screws should be placed on either side of the midline within 2 mm of the site of implantation.
? TROUBLESHOOTING

34| Use the forceps and microspatula to properly anchor the screws into the skull (**Fig. 5f**).

35| Move the drill above the injection site. Drill a hole that penetrates the skull, but not the dura.
▲ **CRITICAL STEP** Drilling through the dura can cause widespread damage, bleeding and inflammation under the skull.

36| Align the infusion pump and lower the injection needle to the dorsal-ventral stereotaxic coordinates of the targeted structure of interest (**Fig. 5g**).
▲ **CRITICAL STEP** A beveled needle can penetrate the dura safely. If a blunt needle is used, the dura should be pierced using a sharp, sterile needle.
? TROUBLESHOOTING

37| By using the microcontroller, infuse the virus at a maximum rate of 100 nl min^{-1}. The volume of the virus will vary depending on the brain structure and serotype of virus used[1,2,4,44–47].
▲ **CRITICAL STEP** All viruses should be kept on ice before use, but special care should be taken with lentiviruses and herpes simplex viruses to prolonged exposure to temperatures above 4 °C.
? TROUBLESHOOTING

38| Once the infusion is complete, allow the injection needle to remain in place for 1 min for every 100 nl of virus infused. Slowly remove the needle from the injection site.

39| Position the cannula holder above the drill hole (**Fig. 5h**). Take care to orient the μ-ILED and other functionalities in the direction suitable for the experiment in order to properly illuminate the opsin-expressing brain region of interest.
▲ **CRITICAL STEP** For injection into rodents, rinse the device with ethanol to sterilize it before injection. For other mammals and primates, it might be necessary to use room-temperature ethylene oxide gas sterilization (http://www.anpro.com/sterilizers/anprolene/indexanprolene.html).
? TROUBLESHOOTING

40| Slowly lower the device into the brain to the desired dorsal-ventral coordinates.

41| By using a syringe needle, slowly apply ACSF to the skull to dissolve the silk-based adhesive. The brain tissue will dissolve the adhesive inside the skull. Wait for at least 10 min to allow for complete dissolution of the adhesive (**Fig. 5i**).
▲ **CRITICAL STEP** The adhesive must be completely dissolved before μ-needle removal. If not, the final placement of the flexible substrates will be affected.

42| Slowly remove the μ-needle. Monitor the position of the flexible substrates to ensure that zero movement occurs. If the adhesive is completely dissolved, the μ-needle will remove with ease and without movement (**Fig. 5j**).

43| Prepare the dental cement as described in Reagent Setup.

44| Use the microspatula to carefully apply a layer of dental cement directly to the point of injection to fully secure the placement of the device (**Fig. 5k**).
▲ **CRITICAL STEP** Allow this layer of cement to completely cure before building the structure of the remaining headcap.

45| Once the initial layer of cement is fully cured, position the PCB-based connector in the desired orientation. Apply a small amount of dental cement to secure it to the base layer (**Fig. 5l**).
▲ **CRITICAL STEP** The orientation of the connector will determine the orientation of the RF headstage antenna.

46| Finish the headstage by completely encapsulating the device-related hardware in dental cement. Be sure to leave the pins of the connector exposed to allow for connections to RF or wired functional generator (**Fig. 5m**,**n**).

47| Use a sterile spatula to detach any scalp skin from the cement. If the skin and the cement are connected, the headstage will be less stable over time.

48| Liberally apply (~1 g) the antibiotic ointment and lidocaine ointment to the entire incision area.

49| Remove the animal from the stereotaxic frame and place it in a clean home cage positioned on top of a heating pad for recovery. The animal should recover rapidly (<15 min), and it can be returned to its home cage once it displays normal,

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Box 2 | Recycling the optoelectronic devices for reuse ● TIMING 2 d

1. After euthanizing the animal, use forceps to forcibly remove the headcap from the surface of the skull. Thoroughly remove any visible biological tissue or other build-up from the sides and bottom of the headcap. Take care not to damage the flexible aspect of the device. Once it has been removed and cleaned, the headcap is ready for dissolution (**Fig. 6a**).
2. Place the headcap into a glass beaker and add methyl methacrylate until the headcap is entirely submerged (**Fig. 6a**). Cover the beaker with tin foil and place in a properly ventilated fume hood overnight.

! CAUTION Methyl methacrylate monomer is a flammable liquid. It may also cause skin irritation; avoid contact with skin, eyes and clothing. Use it with adequate ventilation.

3. After overnight incubation, the dental cement should be completely dissolved. The device and bone screws should be clearly visible in the beaker (**Fig. 6c**). While you are wearing gloves and using forceps, remove these items from the beaker. The screws can be cleaned, sterilized and used for another surgery. The device will normally become inactive after dissolution of the headcap due to concurrent dissolution of the adhesive connecting the ACF cable to both the device and the PCB (**Fig. 6d**,**e**). Before proceeding to Step 4, use a multimeter to check that all μ-ILEDs still function properly (**Fig. 6f**). If any of the μ-ILEDs have electrical failure, discard the device and begin the fabrication process anew.

? TROUBLESHOOTING

4. Discard the original PCB and reassemble the device as described in Step 17 (**Fig. 2g**,**h**).
5. Apply PDMS to the entire device construction to provide passivation and waterproofing for operation *in vivo* as in Step 19 (**Fig. 2i**).
6. Attach the μ-needle using the silk adhesive and test the device to ensure proper electrical connection as done previously in Steps 18 and 20 of the PROCEDURE (**Figs. 2j–l** and **6g**,**h**). The device is now ready for injection into a fresh animal starting at Step 24 of the PROCEDURE.

awake locomotor behaviors. Because the only exposed portions of the devices are the metal connector pins, the animals can be group-housed for the duration of their experimental lifetime. However, if one chooses, the pins can be capped by using a dummy connector.

50| The μ-needle should be clear of any components of the device. Remove the adaptor from the cannula holder and sterilize it for reuse (**Fig. 5o**) as described in **Box 2** and **Figure 6**.

Preparation for behavioral testing ● TIMING ~5 d

51| House mice until the time at which proteins would be expected to be expressed. The choice of viral expression system will determine the expression time, and thus the wait time, before behavioral experimentation. This duration will vary from 1 to 6 weeks. For adeno-associated viral expression, typical wait times are 2–3 weeks for expression at cell bodies[1,2,9,11,44].

52| At least 5 d before experimentation, handle the mice to acclimate them to manipulation. In particular, connectors should be fitted to the headstage, and any areas of the body that will be injected should be gently touched in order to

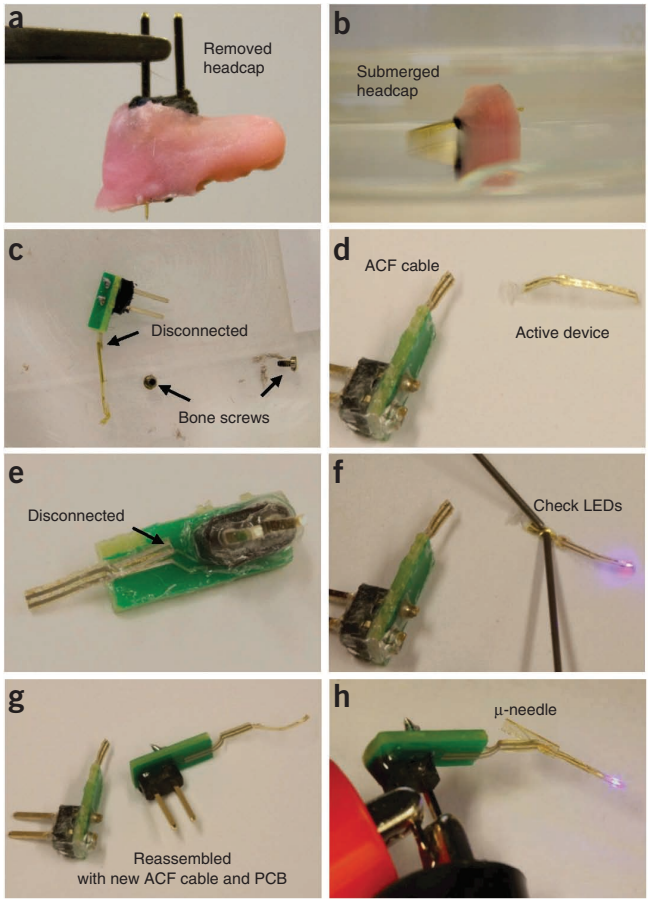

Figure 6 | μ-ILED device recycling and re-fabrication for subsequent use. (**a**) The same headcap from **Figure 5**, after being removed from the animal postmortem and cleaned of biological material. (**b**) The headcap should then be fully submerged in methyl methacrylate. (**c**) After overnight incubation in the stabilized methyl methacrylate monomer, the PCB, connector, μ-ILED device and bone screws will be freely available in the solution. (**d**,**e**) Both connections with the ACF cable will also dissolve, rendering the device inactive. (**f**) The device should be checked for reuse. If it is nonfunctional, the device should be discarded and a new device should be fabricated. (**g**) The working device is reassembled with new ACF cable and PCB. (**h**) The electrical connection through the new ACF cable should be checked after the device is attached to the injection μ-needle again.

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habituate the animal to the manipulations that they will experience during behavioral testing (e.g., the nape of the neck for s.c., the abdomen for i.p., etc.).

53| Also habituate the animals to the method of powering the devices. If a wired approach is used, animals should be connected to the wires and allowed to explore a home-cage environment for at least 20 min once a day for 3 d. If a wireless approach is used, the RF headstage antenna should be connected in the home cage for the same duration.

▲ **CRITICAL STEP** To eliminate locomotor confounds, the animals must be habituated to carrying the added weight of the antennas.

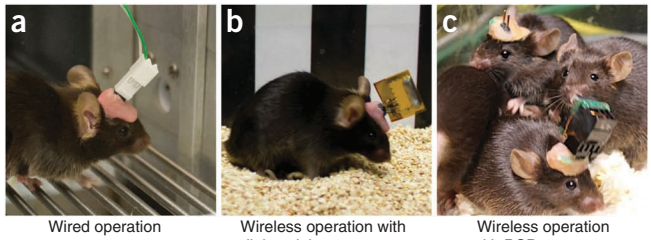

Figure 7 | Expected results after viral and device injection. Once a device is injected, the standard connection allows for temporary coupling multiple means of powering in a variety of behavioral assays. (**a**) A mouse connected for wired powering in a standard operant behavioral chamber. (**b**) The same mouse prepared for wireless powering with the lightweight, flexible power scavenger in a conditioned place preference environment. (**c**) Two mice with implanted devices among cage mates. The mouse in the foreground has a PCB-style RF scavenger for powering in a home-cage environment. All procedures were approved by the Animal Care and Use Committee of Washington University and conformed to US National Institutes of Health guidelines.

54| Perform wired (option A) or wireless (option B) behavioral tests.

▲ **CRITICAL STEP** The timing of the photostimulation will depend on the parameters and goals of the study, but the pulse-generation options are well suited for a wide variety of approaches. Any traditional function generator can either power the devices or provide TTL input into the RF signal generator to modulate the pulse width and frequency of the light pulses. If the photostimulation needs to be contingent on the animal's behavior, this can be achieved by using live video tracking connected to a TTL output (Noldus Ethovision 9.0 with trial and hardware control and I/O box or other) or triggering a TTL signal from IR beam breaks (Med Associates or other).

(A) Wired optogenetic behavioral control ● TIMING variable

(i) Once habituated to the connecting cables, power the devices by using a traditional function generator. The pulse width and frequency of photostimulation should be determined and based on physiologically relevant conditions[7,10,38–40]. The timing of the photostimulation will greatly depend on the parameters and goals of the study.

(ii) Connect the cable to the function generator by using the BNC-to–banana plug adaptor.

(iii) Route the cabling. For most wired behaviors, the best approach is to route the cabling through an electrical rotary joint as described previously[42].

(iv) Carefully scruff the mouse and connect the free end of the cable to the headstage of the animal. Place the animal in the behavioral apparatus (**Fig. 7a**) and perform the desired behavioral test.

(v) After the behavioral test, scruff the animal and remove the cable from the headstage.

(B) Wireless optogenetic behavioral control ● TIMING variable

(i) Connect the function generator to the RF signal generator, the RF signal generator to the RF power amplifier and the RF power amplifier to the panel antenna. Be sure that the power supply is connected to the heat sink (see Equipment Setup) to avoid damaging the amplifier. Connect the power supply to the RF amplifier. It is reasonable to assume that wireless operation of the devices would be compatible with any behavioral assay in which the behavioral apparatus itself does not interfere with the RF signal, but care should be taken to ensure proper powering of the devices in every behavioral apparatus used.

▲ **CRITICAL STEP** To avoid unnecessary powering of the devices, do not engage the power supply until the beginning of the behavioral session.

? TROUBLESHOOTING.

(ii) Carefully scruff the animal(s) and attach the RF scavenging antenna(s). Place the animal(s) in the behavioral apparatus and perform the behavioral test (**Fig. 7b**,**c**).

(iii) After the behavioral test, carefully scruff the animals and remove the scavenging antenna.

(iv) After the final behavioral session, if desired, kill the animals and use their tissue for any manner of post-mortem evaluation[11].

? TROUBLESHOOTING.

? TROUBLESHOOTING

Troubleshooting advice can be found in **Table 1**.

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TABLE 1 | Troubleshooting table.

Step	Problem	Possible reason	Possible solution
14	Failure of the electrical device	Nonuniform coating of BCB passivation on μ-ILED	Spin-coat the BCB again and re-do the process
17, **Box 2** (step 3)	Failure of the ACF cable bonding	Insufficient temperature and/or pressure when bonding	Be certain to apply enough pressure (~1 or 2 MPa) and temperature (150 °C)
24	Device is damaged before or during mounting for surgery	The electrical components are grasped by the cannula holder adaptor	Only grasp the upper, exposed portion of the μ-needle. No portion of the electrical components should be grasped with the cannula holder adaptor. Be careful to apply appropriate pressure so as to avoid crushing the μ-needle
25, 32	Inaccurate tissue targeting	Improperly aligned skull	Follow the instructions provided for your laboratory's stereotaxic equipment for ear bar placement and skull leveling. Species-specific ear bars may be required. If the suture lines for identifying bregma/lambda on the skull are difficult to visualize, a surgical microscope and/or treating the skull with hydrogen peroxide can be a helpful addition. It may be necessary to wait longer after penetration before removal of the μ-needle
33, 36	Headcap falls off in the home cage or during behavior	Insufficient anchoring or cement application	Dental cement is ideal for reusablility, but other adhesives such as VLC one-step adhesive and C&B Metabond offer stronger bonds to the screws and skull surface. Although two screws are normally sufficient, larger PCB interfaces may require more anchor screws
37, 54B(iv)	Little or no expression of the viral construct	Improper handling and injection of the virus	Thaw virus as close in time to the surgery as possible. Keep the virus on ice until the time of injection. Depending on the promoter that drives expression, some systems will not provide sufficient opsin expression. This should be empirically determined before any attempted behavioral experiments
39	Dissolution of silk adhesive during sterilization	Too much exposure to ethanol	Most aqueous solutions will dissolve the silk adhesive. Do not soak the devices in ethanol, as this will promote dissolution of the adhesive. In our experience, rinsing the device with ethanol for ~15 s is sufficient to prevent widespread inflammation in mouse brain tissue[11]. Care should be taken to use appropriate sterilization techniques in higher mammals to avoid immune responses, such as ethylene oxide gas sterilization
54B(i)	Insufficient RF power is reaching the headstage scavenger	The behavioral apparatus is between the panel antenna and the scavenger (in an enclosure)	Connect a freestanding device to the scavenging headstage and place it in the behavioral enclosure. Using the gain on the RF signal generator adjust the RF power until sufficient power is available to power the μ-ILEDs to the desired light output. If problems powering the device persist, consider reducing the distance between the panel antenna and the scavenger
Box 2	Headcap does not fully dissolve	Insoluble dental cement was used, insufficient solvent was used, or there was limited duration of exposure to the solvent	Some dental cements will not dissolve in methyl methacrylate. Consider using the exact recommended cements where possible. If using Jet denture repair, add more solvent or wait longer for the cap to dissolve

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● TIMING

Steps 1–22, preparation of μ-ILEDs, deterministic device assembly and fabrication of wireless power harvester: 9 d
Steps 23–50, viral and device injection: 1 d for procedure, 1–6 weeks for expression
Steps 51–54, behavioral experimentation: 5 d for preparation, variable depending on the experiment
Box 1, combining multifunctional sensors and optoelectronics: 1–4 d depending on modalities
Box 2, rescuing the optoelectronics for re-use: 2 d

ANTICIPATED RESULTS

After successful fabrication and implementation of μ-ILEDs, wireless manipulation of intact mammalian neural circuitry is possible. Although our group has demonstrated the feasibility of using these devices to investigate reward-related and anxiety-like behaviors[11], nearly any behavioral assay should be accessible with these devices. In particular, behaviors that require complete freedom of movement such as social interactions, wheel running and home-cage behaviors are now possible (**Fig. 7c**). Users can expect injected devices to function for many months, as we have tested devices up to 6 months after injection and observed that the devices retain their operational functionality[11]. Furthermore, we show here that these devices can be reused in other animals to avoid the lengthy process of remaking new devices for each experiment. Depending on the nature of the experiment, a single panel antenna can power numerous headstage devices. If the stimulation parameters are the same across animals, this approach can greatly increase the experimental throughput for a variety of behavioral assays. Although not presented in detail here, a reasonable extension of this protocol would be to incorporate other existing microscale sensors such as those for pH, blood oxygen, glucose levels or neurochemical detection onto μ-ILED devices[48,49]. In sum, the protocol described here should provide researchers with the ability to use optogenetics to wirelessly control and study intact mammalian neural circuitry in a minimally invasive manner.

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Note: Any Supplementary Information and Source Data files are available in the online version of the paper.

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AN IMPROVED RECTENNA FOR WIRELESS POWER TRANSMISSION FOR UNMANNED AIR VEHICLES

by

Chun-Yi Liu

September 2011

Thesis Advisor: David C. Jenn
Second Reader: Ric Romero

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13. ABSTRACT (maximum 200 words)

This thesis continues an NPS project related to wireless power transmission for micro air vehicles (MAVs). The conversion of radio-frequency (rf) power into usable direct-current (dc) power is performed by a rectifying antenna, or rectenna. The emphasis of this thesis is the simulation and experimental study of various rectenna designs to determine which best provides high efficiency, stable output power, and lightweight design.

The analysis of rectenna design focuses on four subsystems: (1) the receiving antenna, (2) the matching sections, (3) the rectification, and (4) the post-rectification filter. Based on the findings of this research, the ultimate rectenna design implements a half-wave dipole antenna that performs full-wave rectification with two diodes. The post-rectification filter is implemented by a capacitor to obtain stable dc power. The final design achieved an efficiency of nearly 66% for input power in the range of 200 mW.

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AN IMPROVED RECTENNA FOR WIRELESS POWER TRANSMISSION FOR UNMANNED AIR VEHICLES

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Captain, Republic of China Army
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Submitted in partial fulfillment of the
requirements for the degree of

MASTER OF SCIENCE IN ELECTRONIC WARFARE SYSTEMS ENGINEERING

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ABSTRACT

This thesis continues an NPS project related to wireless power transmission for micro air vehicles (MAVs). The conversion of radio-frequency (rf) power into usable direct-current (dc) power is performed by a rectifying antenna, or rectenna. The emphasis of this thesis is the simulation and experimental study of various rectenna designs to determine which best provides high efficiency, stable output power, and lightweight design.

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LIST OF ACRONYMS AND ABBREVIATIONS

ADS	Advanced Design System
BPF	Bandpass Filter
CST	Computer Simulation Technology
CW	Continuous Wave
dc	Direct Current
DARPA	Defense Advanced Research Project Agency
FGCPW	Finite-width Ground Coplanar Waveguide
LPF	Lowpass Filter
LTSA	Linear Tapered Slot Antenna
MAV	Micro Air Vehicle
rf	Radio Frequency
SL	Slot Line
SPS	Solar Power Satellite
S-Parameters	Scattering Parameters
SSP	Space Solar Power
UAVs	Unmanned Air Vehicles
WPT	Wireless Power Transmission

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I. INTRODUCTION

A. Background

The concept of wireless power transmission (WPT) dates back to the days of Heinrich Hertz and Nikola Tesla [1], who discovered that energy could be transported by electromagnetic waves in free space. Tesla considered the use of wireless power transmission employing low-frequency transmission, sustained by the earth's natural electromagnetic resonance. His concept is important to later studies of electromagnetic-wave propagation. There has been intense interest in WPT recently for a number of applications. These include remote propulsion of vehicles [2], transmission of solar power from space [3] and wireless battery charging [4]. A related area is energy harvesting, where stray electromagnetic fields from the many systems in the environment are collected and used as a free energy source [5].

Micro-air-vehicle (MAV) propulsion is the main application of interest in this research. MAVs are a category of unmanned air vehicles (UAVs) being developed around the world. The definition of an MAV, according to the Defense Advanced Research Project Agency (DARPA), is a fully functional UAV no larger than 15 cm in length, width, and height [6]. The main reason for using wireless power transmission for military ground-surveillance applications is to achieve unlimited duration of flight without requiring onboard fuel supplies. One application that employs a MAV remotely powered by a small ground station is illustrated in Figure 1.

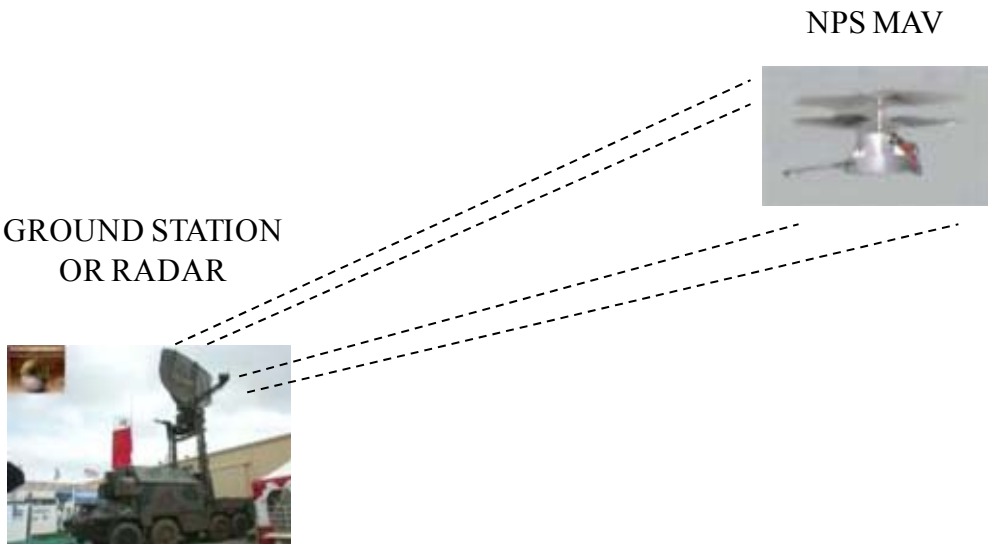

Figure 1. Conceptual powering of an MAV by a ground station (From [6]).

In this thesis, a rectifying antenna (rectenna) system based on studies performed by Tan [6] and Toh [7] was re-evaluated, simulated, and refined. This study also looks at a new antenna design and a full-wave rectifier concept. In order to verify and validate the design of the rectenna model, many tradeoffs were made. Different rectenna designs were investigated, and an improved design was implemented in hardware.

The measured efficiency of the previous NPS rectenna was 26% to 37% [7]. This research changes the circuit design to yield better overall performance. The recommended design changes are shown to yield higher efficiency and lighter weight. The claimed theoretical efficiency of the rectifier designed by Toh [7] was approximately 26-37%, which was insufficient to power a micro air vehicle designed by Tsolis [8].

B. OBJECTIVE

The purpose of this thesis is to verify and analyze the different designs of the rectenna system and make improvements. Since the primary application considered is UAV propulsion, an important requirement is to minimize the antenna and rectifier circuit's size and weight. A new dipole design is presented that is lightweight and compact. A full-wave rectifier circuit is proposed, which increases the output power over that of the half-wave rectifier design.

C. THESIS OUTLINE

This thesis is divided into six chapters. The next five are organized as follows:

Chapter II introduces the development of WPT from the early 1950s through today. References to various WPT studies and applications are also covered in this chapter. An overview of related thesis projects that were conducted at the Naval Postgraduate School is presented.

Chapter III focuses on three different rectenna designs, the components of a rectenna circuit, and evaluation of various circuit designs by using Advanced Design System (ADS) 2009 software from Agilent Technologies. Analysis of the dipole antenna dimensions, Schottky diode, and substrate material are covered in this chapter. Chapter III also discusses the effects of pre- and post-rectification filtering. The design of a rectenna with and without filters is documented, and a comparative study of the performance of the pre-, post-, and no-filter designs analyzed.

Chapter IV discusses the impedance of the rectenna and dipole-antenna array.

Chapter V summarizes the findings of this research and presents conclusions and recommendations for future research.

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II. BACKGROUND

A. EARLY EXPERIMENTS WITH WPT

The concept of wireless power transmission (WPT) began in the 1900s. The objective of WPT is to transfer electrical power through space without the need for cables. Conventional communication systems such as radio and cell phones only transmit small amounts of energy, but WPT must transfer a significant amount of power in order to charge a battery or propel vehicles. A generic WPT block diagram is shown in Figure 2. A key component in the receiving system is the rectifying antenna, or rectenna.

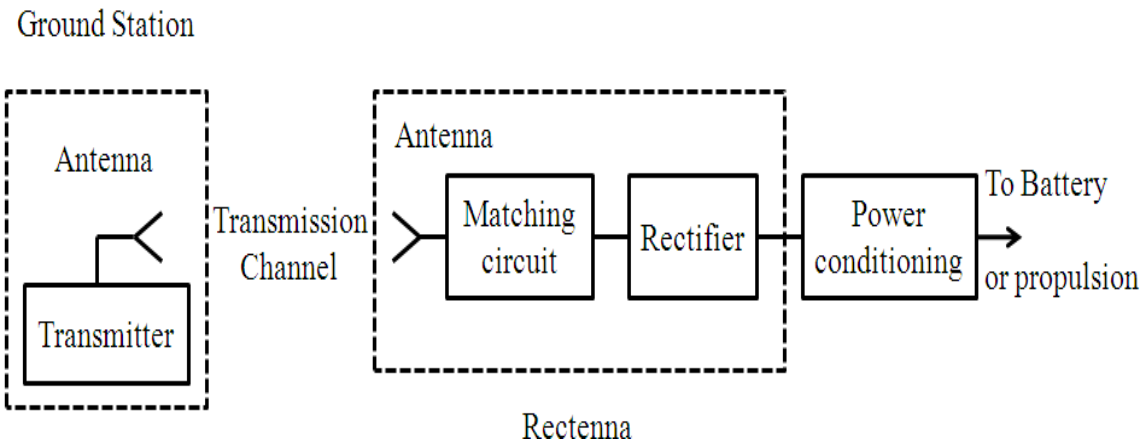

Figure 2. Generic WPT-system block diagram.

Brown conducted wireless power-transmission research in 1963 [1]. In his article, Brown wrote a description of the development of WPT and the milestones along the way. Previously, Tesla carried out numerous experiments to transmit power without a wire at Colorado Springs in 1899 (Figure 3). A gigantic coil was built and fed with 300 kW of low-frequency power in order to produce very long and visible discharges from the sphere, but there is no record of whether any significant amount of power was collected at a distant location. Although Tesla's experiments ended due to lack of land and funding, his concept of efficient, wireless power transmission was important to future research.

Figure 3. Experiment with radio waves for power transmission (From [1]).

In the late 1930s, a further advance in WPT occurred with the invention of the klystron tube, which converted microwave power into dc power by using microwave power tube at one end and dc diode tubes at the other. Advances in the microwave-cavity magnetron led to higher efficiency for WPT applications during World War II.

In the 1950s, two advances were enabled by the invention of the amplifier tube, which created a larger amount of transmitting power [9] to drive an electromagnetic beam, and the focusing of electromagnetic power into a beam for high efficiencies [10]. In May 1963, Raytheon demonstrated the first microwave-power transmission system, which converted 400 W of CW power at the transmitter to 100 W of dc power to drive a motor.

In October 1964, a demonstration of microwave-powered helicopter flight up to 60 ft. above a transmitting antenna was presented (Figure 4). In November of that year, a nonstop, 10-hour hover was demonstrated.

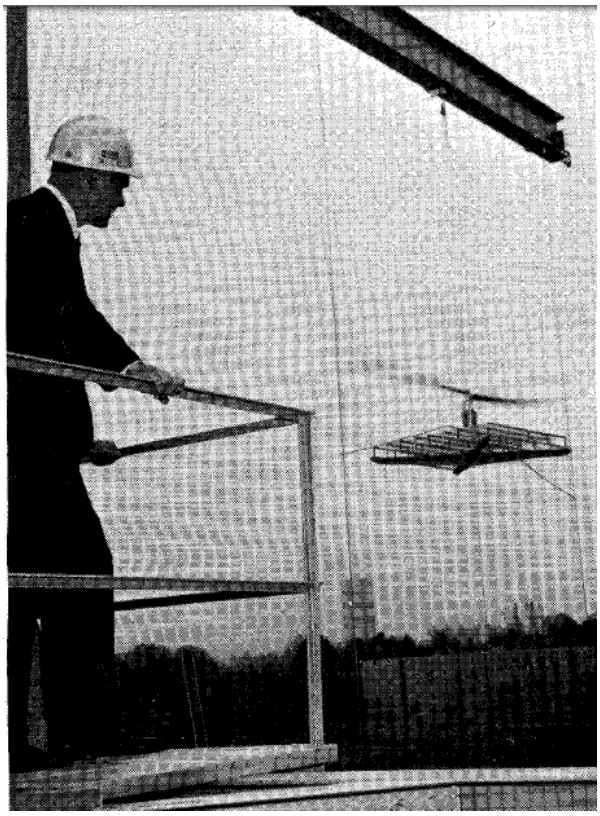

Figure 4. A helicopter powered by WPT (From [1]).

The concept of a solar-power satellite (SPS) introduced in the 1960s also relied on WPT, as illustrated in Figure 5. In the SPS concept, solar energy collected in orbit is converted into microwave power for transmission to a large antenna on ground. With oil consumption increasing over the past years, WPT technology has become an alternative source of energy by transmitting collected solar power from satellites to an earth rectenna station. Solar energy transmitted by WPT is environmentally clean and available 24 hours a day from space by using satellites as collection stations. Although the SPS program ended in 1980, it redirected the design of the transmitter antenna to an active, phased array made from a large number of microwave generators instead of super power tubes.

For ground-based arrays, the low-cost, microwave-oven magnetron could be used directly in the SPS. Pacific Gas and Electric (PG&E) has recently funded a study to design a commercial SPS system [3].

Figure 5. Satellite solar-power model (From [14]).

In the 1990s, the advancement of WPT related to the use of solid-state silicon-based PN junction diodes, which have a high turn-on voltage, and new GaAs Schottky diodes that have fast switching capability for the high-frequency rectification processes. Semiconductor devices were became suitable for achievement of high-rectification efficiency in WPT applications. The first fuel-less plane powered by microwave was produced by Joseph in September 1987 [2]. An array of antennas was used to transmit microwaves onto the plane disc with a dual-polarized rectenna. The airplane flew on beamed microwave for 20 minutes.

B. RECENT DEVELOPMENTS IN WPT

Recently there has been increased activity in WPT. In a September 1999 article by Youn [11], a WPT system achieving a single rectenna conversion efficiency of 75.6% and an overall system efficiency of 33% are described.

Hagerty [12] proposed an array of 64 circularly polarized spiral elements and used a broadband-antenna array to convert more microwave power into dc power. The achievement of 20% efficiency over a frequency range of 2–18 GHz was reported.

A December 2002 survey article on space solar-power (SSP) program research was written by McSpadden [13]. McSpadden gave insight on achieving practical SSP and emphasized three critical components: the transmitter, beam control, and rectennas. Although it was acknowledged that the technology was immature, a strategic roadmap was highlighted for future investigation.

In [14], Goel proposed that a solar-power satellite could be placed in halo orbit to be fixed over the earth's midnight as the earth rotates. The proposed orbit, known as the trans-earth Lagrange point, obviates the need for a rotating joint between the solar array and microwave transmitter, making a simpler design possible.

C. RECENT RESEARCH IN RECTENNAS

In [15], a dual-frequency. printed-diode rectenna system using a GaAs Schottky diode that could achieve 84.4% and 82.7% efficiency at 2.45 and 5.8 GHz, respectively, is reported. The dual-frequency rectenna system is illustrated in Figure 6.

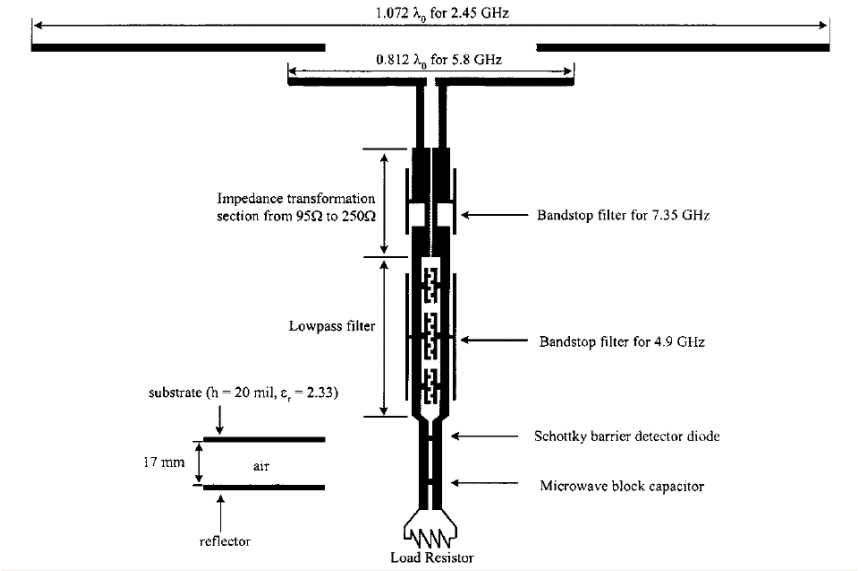

Figure 6. Circuit configuration of the dual-frequency rectenna (From [15]).

Reference [16] by Strassner contains an article on a 5.8 GHz, circularly polarized rectenna design and array. Strassner demonstrated the feasibility of a rectenna array scaled 3 feet wide by 2 feet long and 4 inches tall, with an efficiency of 82% and output power of 7.6 W.

Reference [17] reports a circular-sector antenna using inset feeding in order to exhibit high reflection coefficients at the second and third harmonics generated by the diode. As illustrated in Figure 7, the rectenna can eliminate the need for a low-pass filter (LPF) placed between the antenna and the diode, as well as achieve a conversion efficiency of 77.8% at 2.4 GHz.

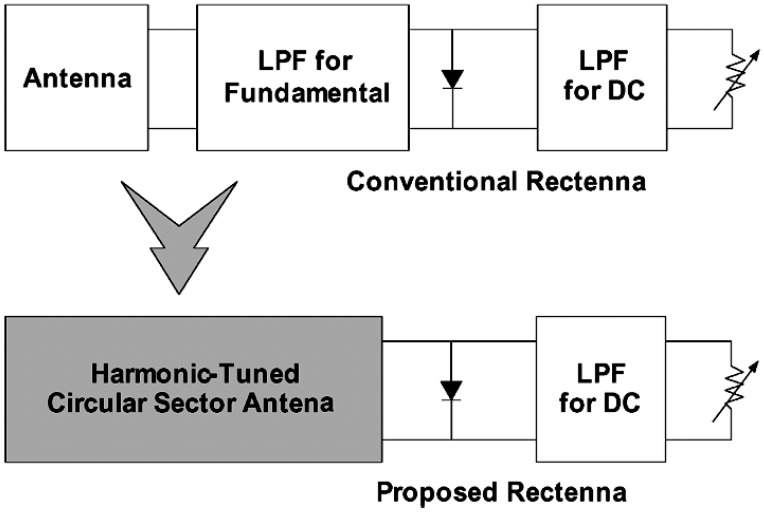

Figure 7. Block diagram of the conventional and the proposed rectenna (From [17]).

In [18], Zbitou reported a monolithic rectifier design with a second diode acting as a variable resistor to achieve an overall efficiency of 65% at 2.54 GHz. He conducted the study using a GaAs, pseudomorphic, high-electron-mobility transistor process to develop a monolithic rectifier to minimize circuit dimensions. The dimension of the rectenna was 1340 μm by 476 μm.

The use of full-wave rectification for rf-to-dc conversion showed that 70.69% rectenna efficiency could be achieved with an input power of 45 mW [19]. The conversion efficiency depends on load resistance, due to the internal resistance of the rectenna system (Figure 8).

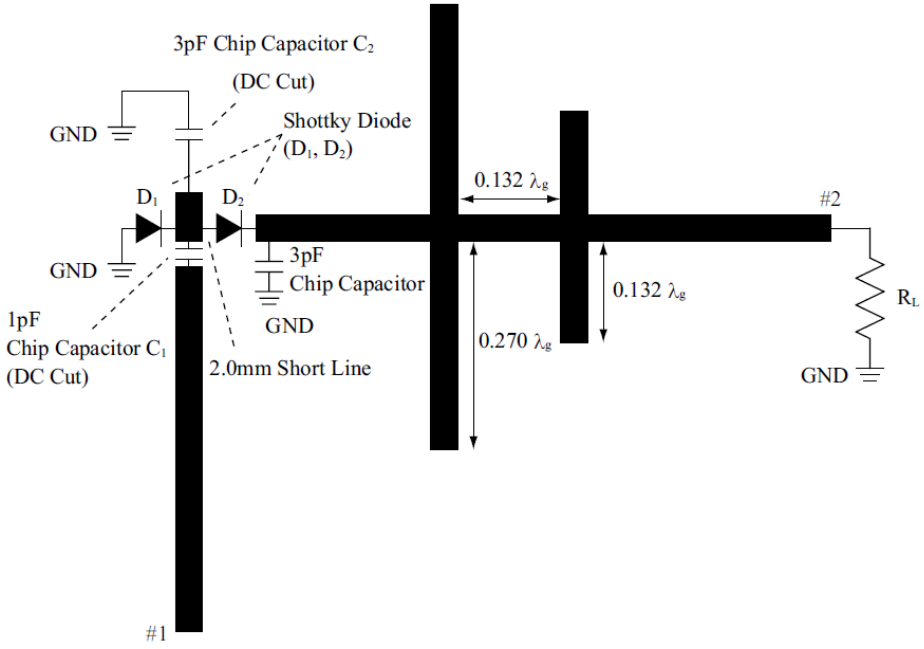

Figure 8. Full-wave rectification (From [19]).

A recent article by Olgun [20] reported that a rectenna system combining a two-stage, zero-bias Schottky diode with a miniature antenna could achieve 70% efficiency at 2.4 GHz. The diodes are in parallel to rf signals, but appear in series for the dc circuit in order to produce doubled voltage. The circuit layout of the rectenna is depicted in Figure 9.

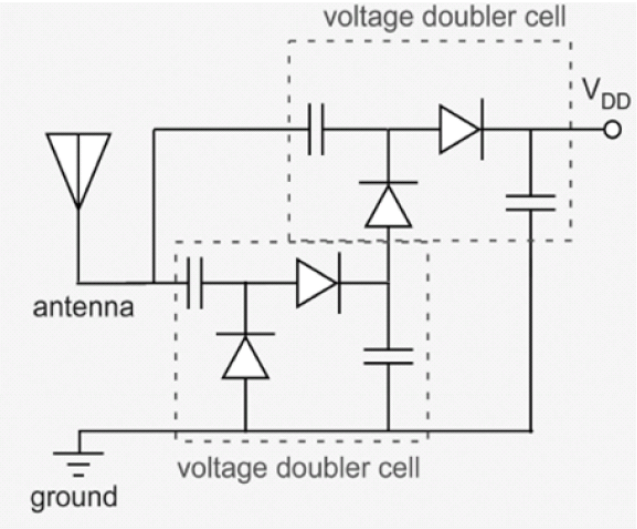

Figure 9. Circuit design of the rectenna (From [20]).

D. PREVIOUS NPS RESEARCH

Students at Naval Postgraduate School have looked into aspects of WPT for MAV propulsion. Vitale [21] utilized a metal semiconductor diode to design a rectification circuit and was able to experimentally determine the scattering parameters (S parameters) of the diode. He highlighted that future research into employing GaAs Schottky diodes for higher rectification efficiency and using a high power source for examining improvement on the MAV antenna design.

Tsolis [8] extended Vitale's work by using Schottky diodes with a patch antenna, but efficiency was verified to be only 7%, due to the mismatch of impedance. Tan [6] investigated various designs of antennas and concluded that a round-patch antenna and sixth-order filters were the better design. He recommended that further analysis into building rectenna to determine the fabrication robustness and conversion efficiency.

Toh [7] improved Tan's design by matching the impedance of the rectenna system. Toh introduced an impedance-matching unit and simulated his design using software. The prototype of the rectenna system was fabricated and tested. The overall efficiency of a single rectenna element varied between 26% and 37%.

E. SUMMARY

In this chapter, the major milestones in the development of WPT were highlighted and a brief history of WPT was covered. Much research has been conducted to determine the feasibility of implementing WPT and its applications, which range from miniaturized versions of remotely powered vehicles to large-scale systems. In the next chapter, rectenna design is addressed in detail.

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III. RECTENNA DESIGN

A. DISCUSSION OF RECTENNA DESIGN

1. General Rectenna Design

A rectifying antenna (rectenna) receives a microwave signal at the antenna and converts it to direct current. It should do this as efficiently as possible and provide a clean, constant, low-ripple voltage. Typically, a rectenna is composed of four components: (1) an antenna, (2) a pre-rectification filter, (3) a rectification diode, and (4) a post-rectification filter [7]. The basic configuration of the previous NPS rectenna systems is depicted in Figure 10.

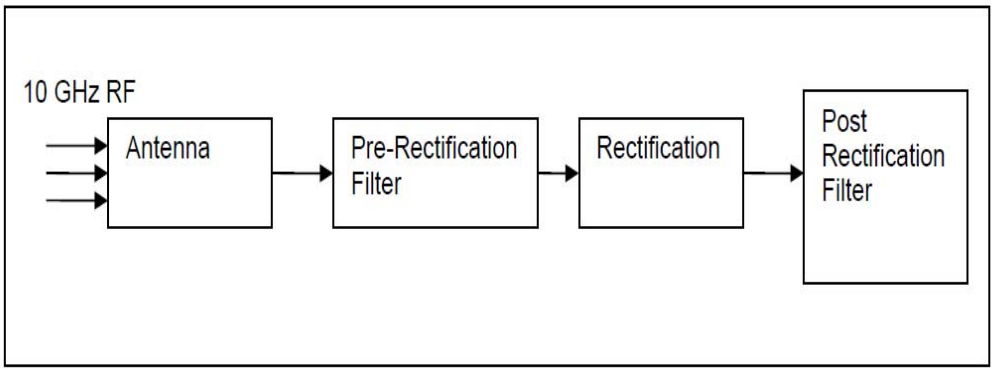

Figure 10. Basic configuration of NPS rectenna system (From [7]).

There are many candidate antenna designs, such as half-wave dipole, horn, parabolic antenna, and microstrip antenna. Some of these can be discarded for this application, due to their weight or size. In reference [8], Tsolis concluded that a circular-patch antenna was preferred, in part due to its capability of reducing the reradiation of harmonics. Figure 11 shows the S_{11} (return loss) response of a circular-patch antenna. Ideally, the return loss of a circular-patch antenna is about 50 dB. In practice, the return loss is likely to be 15 dB to 20 dB. (Return loss is a positive quantity equal to the negative of S_{11} in dB.)

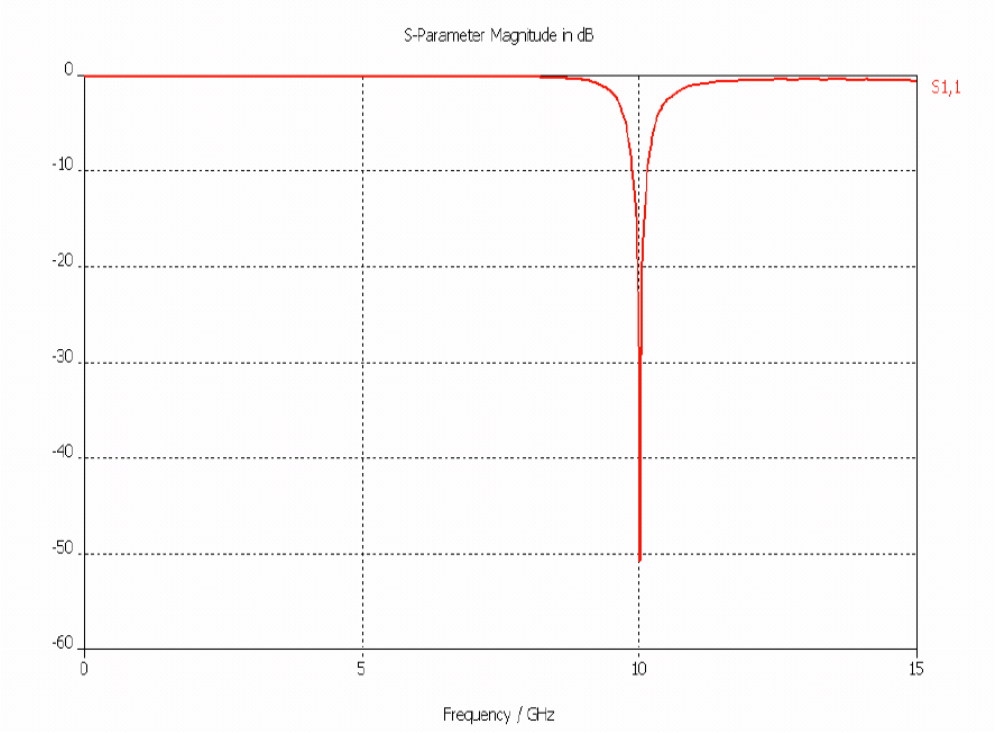

Figure 11. S_{11} frequency response of circular-patch antenna design (From [7]).

The antenna feeds directly into a pre-rectification filter. It allows the 10 GHz operating frequency to pass and prevents interference signals and re-radiation of higher-order harmonics generated by the diode. It also helps provide a constant load impedance for the antenna. Tan [6] selected a sixth-order LPF instead of a fourth-order LPF, due to its better response at 10 GHz. Figure 12 shows that the S_{12} (insertion loss) is 0.078 dB (insertion loss is the negative of S_{12} in dB.)

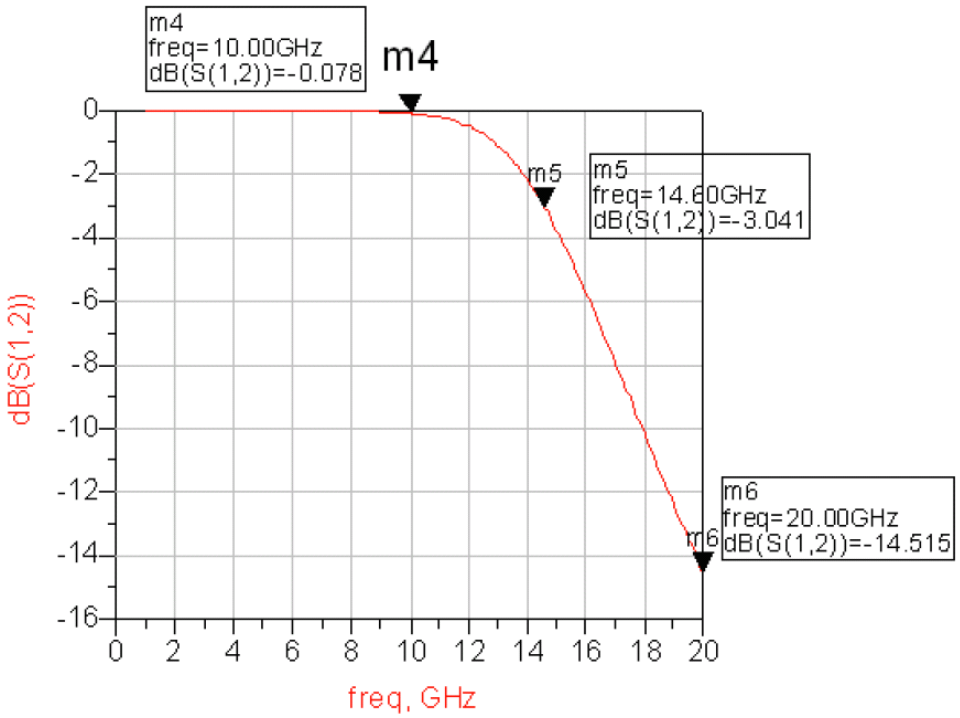

Figure 12. Simulated response of sixth-order low pass filter using ADS software (From [7]).

Rectification is the key function of the rectenna: converting microwave rf energy into dc power. In reference [8], the Schottky diode was used as the rectifying device, due to a high switching capacity that enables it to follow a high-frequency input signal. A commercial Avago HSMS8101 was used in the NPS rectenna system [7], [8]. The S_{21} response of the Schottky diode is depicted in Figure 13. The value of S_{21} is about -0.5 dB at 10 GHz [7].

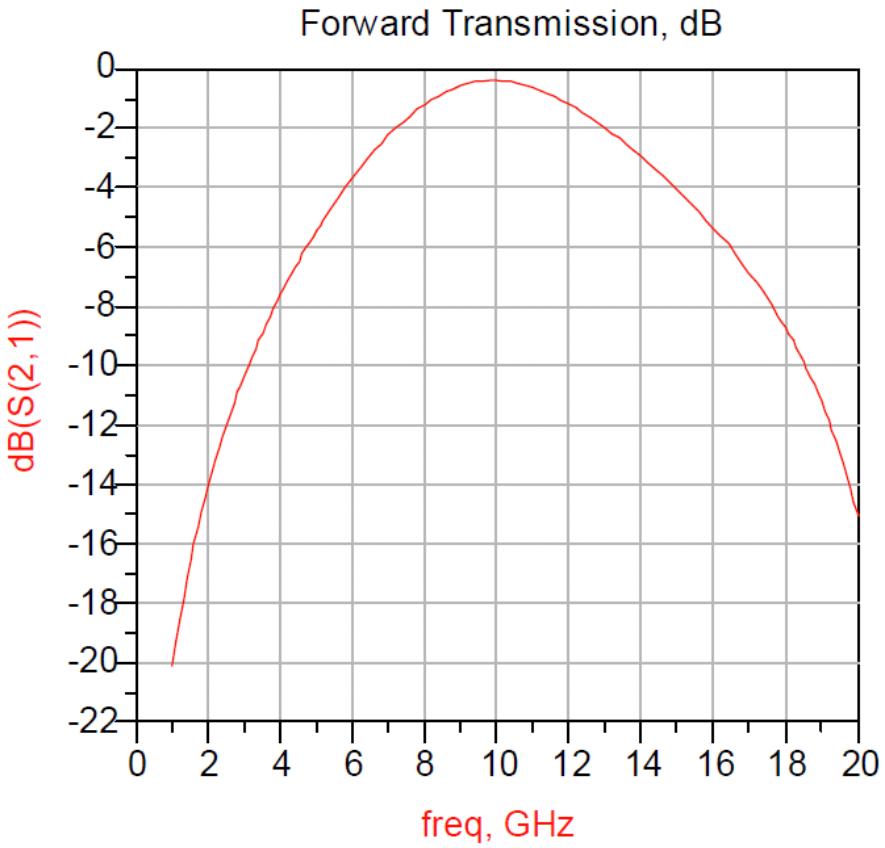

Figure 13. Simulated response of Schottky diode using ADS software (From [7]).

A post-rectification filter is used to extract the dc component and produce a smooth transient for output dc power. Toh [7] developed a post-rectification filter by simply using a 47 pF capacitor in the circuit design.

Tsolis [8] studied the MAV motor prototype DMMK06-10 and measured data for the motor. Toh [7] confirmed the characteristics of the motor given in Table 1. In order to hover the weight of 10 g with 0.952 W, Toh concluded that at least 22 rectenna elements are needed, with each rectenna producing a maximum power of 75 mW. He also implemented a half-wave rectenna and measured the maximum power produced by a single rectenna to be 41.15 mW. In order to hover the MAV, the present research needed to produce more dc power from each rectenna than the previous half-wave rectenna and design a lighter-weight rectenna circuit.

DMMK06-10 Motor's characteristics							
Weight (g)	net	5	10	15	20	25	30
Voltage (V)	5.5	7	8.1	9.9	11.5	14	16
Current (A)	0.1	0.136	0.17	0.22	0.264	0.319	0.357
Resistance (Ω)	54.9	51.4	47.4	45	43.5	43.8	44.8
Power (W)	0.55	0.952	1.3817	2.1780	3.0360	4.4660	5.7120
Round Per Minute	1200	1440	1620	1740	1770	1800	2100

Table 1. Measured data for the MAV motor (From [7]).

2. Antenna Design

In order to obtain a lightweight rectenna design for MAV applications, we discuss possible changes or removal of components of the rectenna. First, we propose using a half-wave dipole antenna instead the original circular-patch antenna to reduce weight. Figure 14 illustrates the structure of the half-wave dipole antenna over a ground plane modeled in CST Microwave Studio. The dipole has a unique design: only one arm is fed by the microstrip line, while the other arm is open. As will be discussed in a later section, this represents the situation that occurs when the dipole is used in a full-wave rectenna.

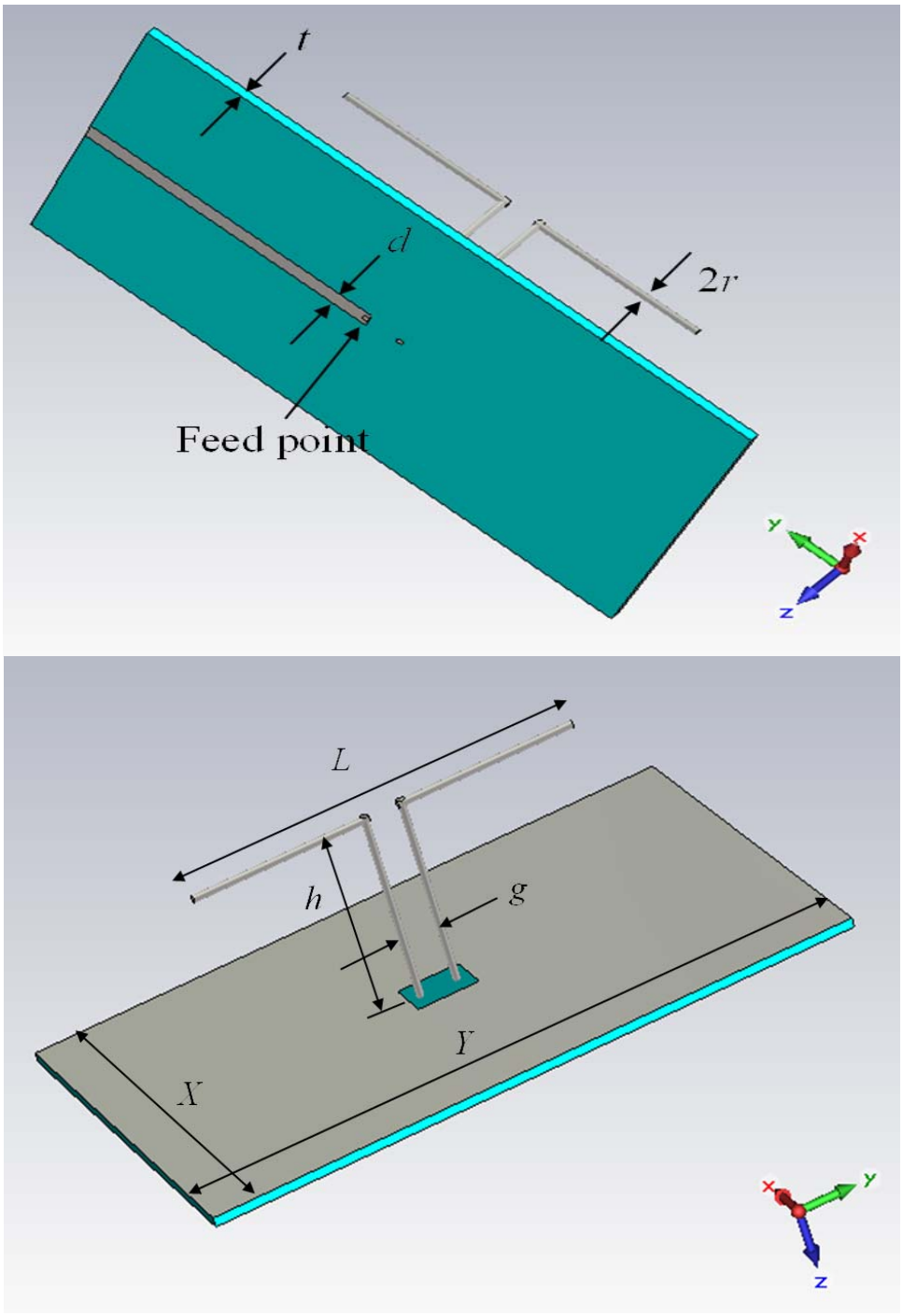

Figure 14. Two views of the half-wave, dipole-antenna structure.

Table 2 lists the design specifications for the data in Figure 15 through 17.

Half-wave dipole's specifications			
Dipole wire radius (r)	0.25	Gap width (g)	1.5
Substrate height (t)	0.508	Dipole height above ground (h)	9
50 ohm line width (d)	1.6829	Sample width (X)	25
Dipole length (L)	25	Sample length (Y)	35

Table 2. Design specifications for the half-wave dipole antenna.

Figure 15 shows the S_{11} response of the half-wave dipole antenna. The return loss is about 23.5 dB at 10 GHz. The impedance of the antenna is 44+j1.6 Ω at 10 GHz as noted on the Smith chart in Figure 16.

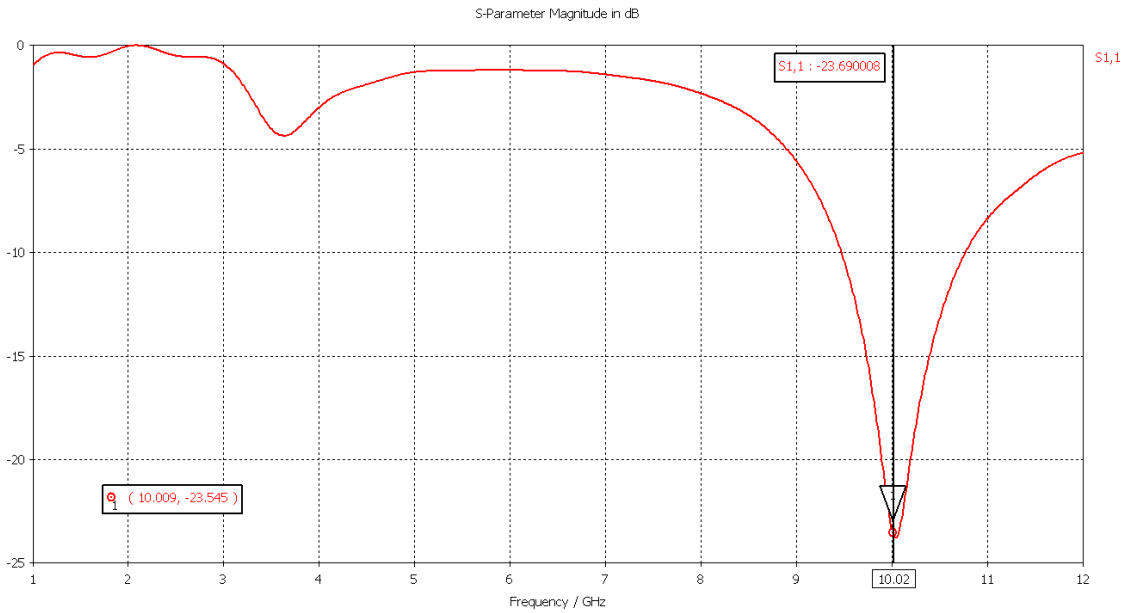

Figure 15. S_{11} frequency response of half-wave, dipole-antenna design.

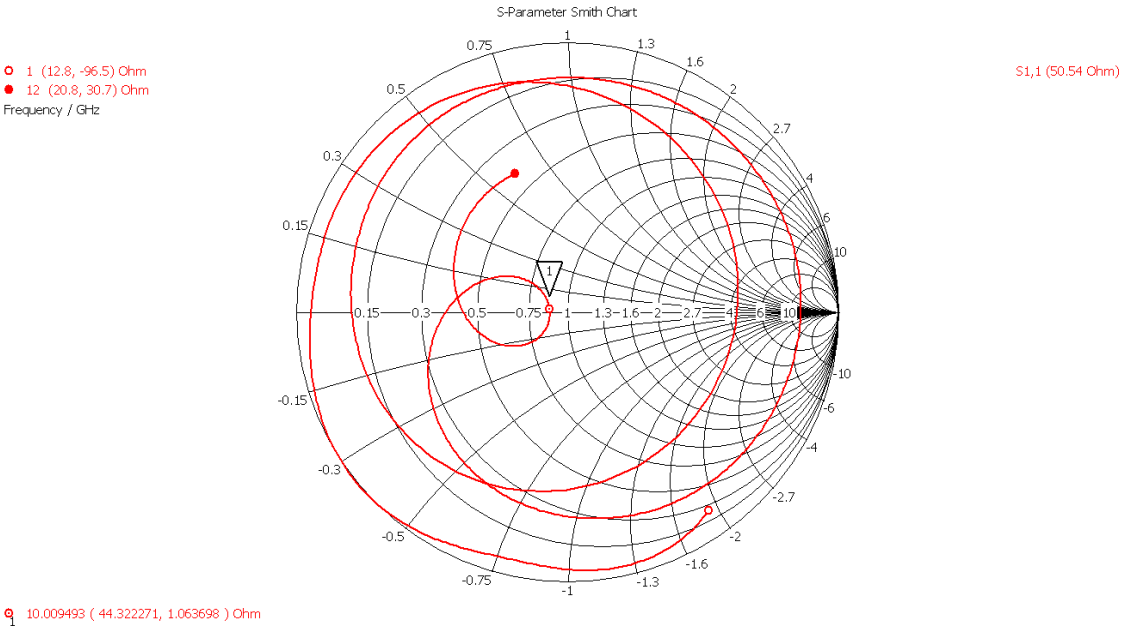

Figure 16. Smith chart of the half-wave, dipole-antenna design**.**

The pattern, gain, and radiation efficiency of the antenna are given in Figure 17. The coordinate system is defined in Figure 14. The pattern plotted is $|E| = \sqrt{|E_\theta|^2 + |E_\phi|^2}$. The pattern gives good coverage over a wide range of angles. The large backlobe is from radiation by the microstrip feed line. It will be shielded in the final design. The return loss and pattern were found to be sensitive to the ground-plane size and dipole height. Smaller ground planes required a higher dipole for a good impedance match. From the simulation, we observe that using a half-wave dipole antenna has acceptable return loss and pattern.

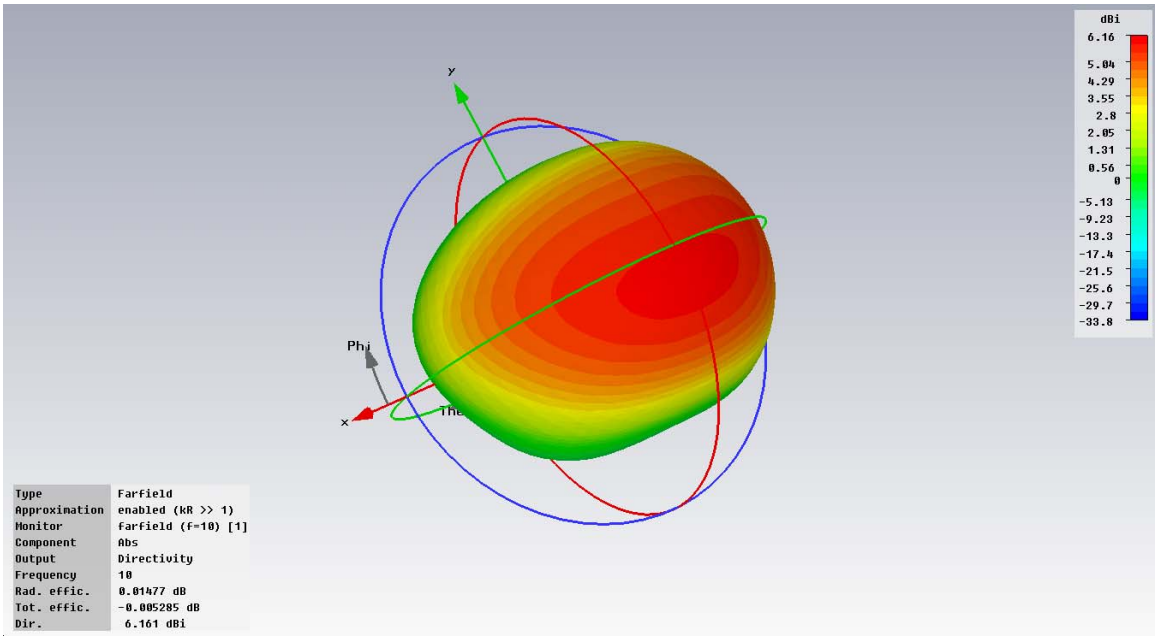

Figure 17. Three-dimensional, far-field radiation pattern of the half-wave, dipole-antenna design.

3. Power Budget

In this section, we introduce the effective area of the transmitting and receiving antennas and calculate the power density in order to obtain the required transmitter power for various distances. Figure 18 shows the geometry.

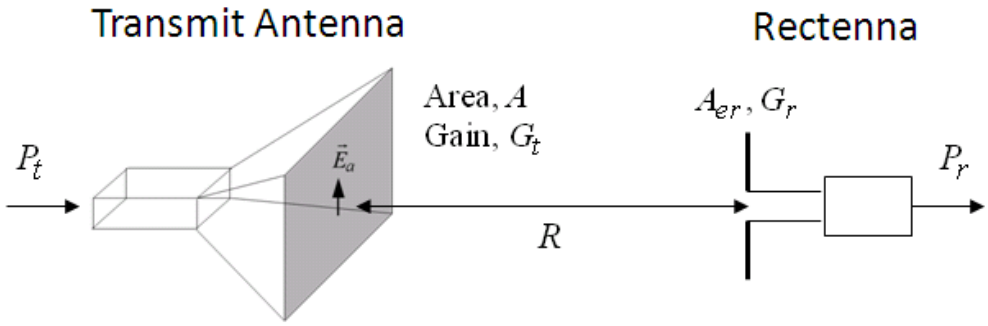

Figure 18. Diagram of the transmit antenna and rectenna.

The effective area of the dipole antenna over a perfect ground plane is calculated using:

$$A_{er} = \frac{G_r \lambda^2}{4\pi} \tag{3.1}$$

where G_r is the gain of the dipole antenna and λ is the wavelength at 10 GHz. From Figure 17, the gain is 6.16 dB = 4.13. Using Eq. (3.1), the effective area of the dipole antenna is 2.96×10^{-4} m^2. The power density W_i transmitted by the horn antenna at distance R can be expressed as

$$W_i = \frac{P_t G_t}{4\pi R^2} = \frac{4\pi P_t \frac{A_{et}}{\lambda^2}}{4\pi R^2} = \frac{P_t A_{et}}{\lambda^2 R^2} \tag{3.2}$$

where P_t is the transmitted power, A_{et} is the effective area of the horn antenna, and R is the distance between transmitting and receiving antennas. The effective area of the horn antenna is the physical area (A) times the efficiency, e_t. Assuming horn dimensions of 3 inches by 4 inches and an efficiency of 0.7, $A_{et} = 0.0054$ m^2, the power P_r received by dipole antenna can be expressed as

$$P_r = W_i A_{er} = \frac{P_t A_{et} A_{er}}{R^2 \lambda^2}. \tag{3.3}$$

Using Equation (3.3), the required transmitting power versus the distance between transmitting and receiving antennas to obtain 200 mW of received power is given in Figure 19. From Figure 19, the transmitted power is 1.35 W for the distance of 30 cm, and the received power will be 200 mW. This analysis assumes the distance R is in the far field of the horn antenna.

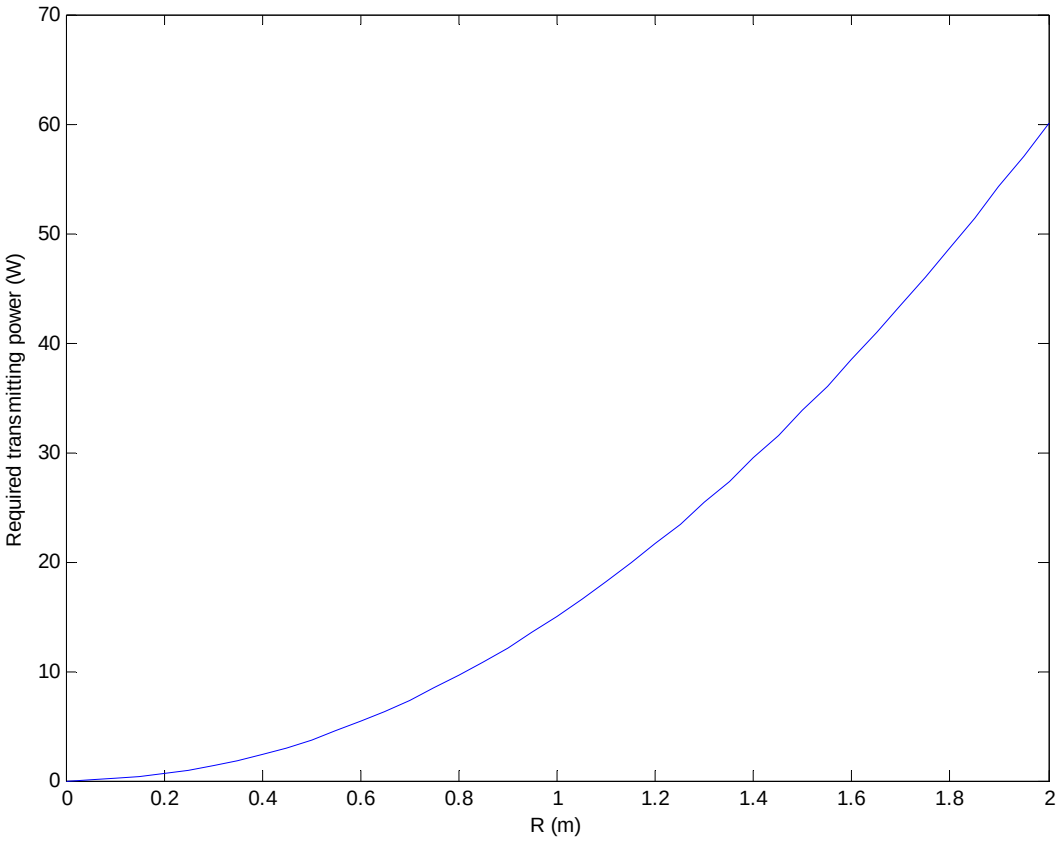

Figure 19. Required transmitting power versus distance with 200 mW received power.

Next, we evaluate the various rectenna configurations by simulating the conversion efficiency and discuss the effect of changing the position of the pre-rectification filter in rectenna design by analyzing the harmonics in the frequency domain.

B. COMPARISON OF DIFFERENT DESIGNS USING ADS

In this section, we simulate the rf-to-dc efficiency of different rectenna designs, such as a half-wave-rectification rectenna, hybrid rectenna, and full-wave rectenna. Agilent ADS2009 is used to evaluate and compare the performance of the various designs. This software provides numerous simulation technologies, covering issues such as electromagnetic-field simulation and frequency and time-domain circuit simulation. To compare performance with the NPS rectenna design, we model the original rectenna circuit of Toh [7] using the HSMS8101 Schottky diode and sixth-order LPF.

1. Half-Wave-Rectification Rectenna Design

The half-wave-rectification rectenna has the advantage of simple circuit layout and ease of fabrication as a microstrip structure [22]. A block diagram of a half-wave-rectification rectenna is depicted in Figure 20.

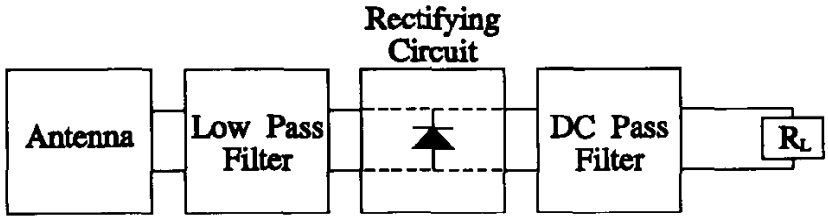

Figure 20. Block diagram of half-wave rectification rectenna (From [22]).

In references [6], [7] a half-wave rectifier design was used. The circuit layout of the rectenna is given in Figure 21.

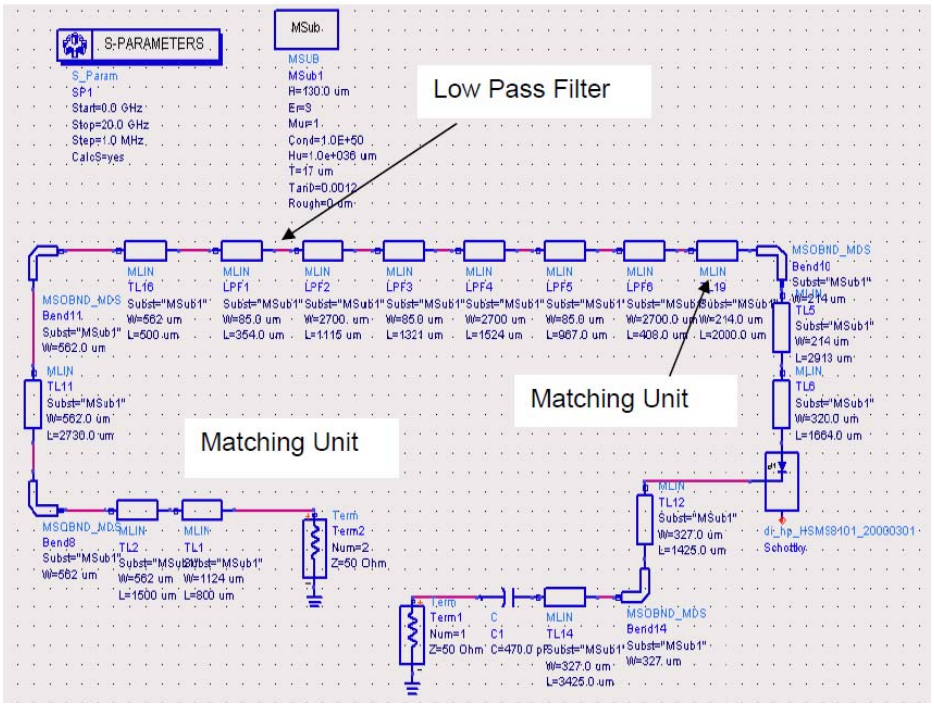

Figure 21. ADS S-parameter circuit model of Toh's rectenna (From [7]).

Toh removed the dc pass filter because it was too small to solder correctly and it was found not to significantly affect output [7]. Omitting this component results in some ripple in the output dc; however, to minimize size and weight, the low-pass filter is possibly a component that can be omitted for some circuit designs. Figures 22 to 24 illustrate three different circuit designs of a half-wave rectenna that will be simulated in ADS. The three half-wave rectifier designs are: (1) with a pre-LPF, (2) with a post-LPF, and (3) without any LPF.

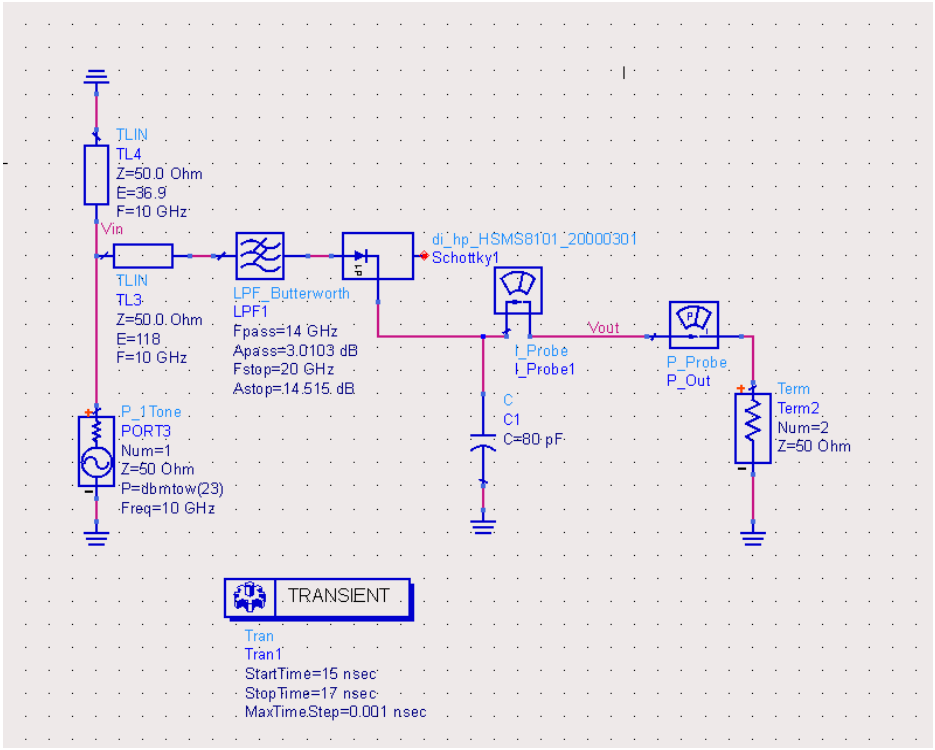

Figure 22. ADS model of a half-wave rectenna design with a pre-LPF.

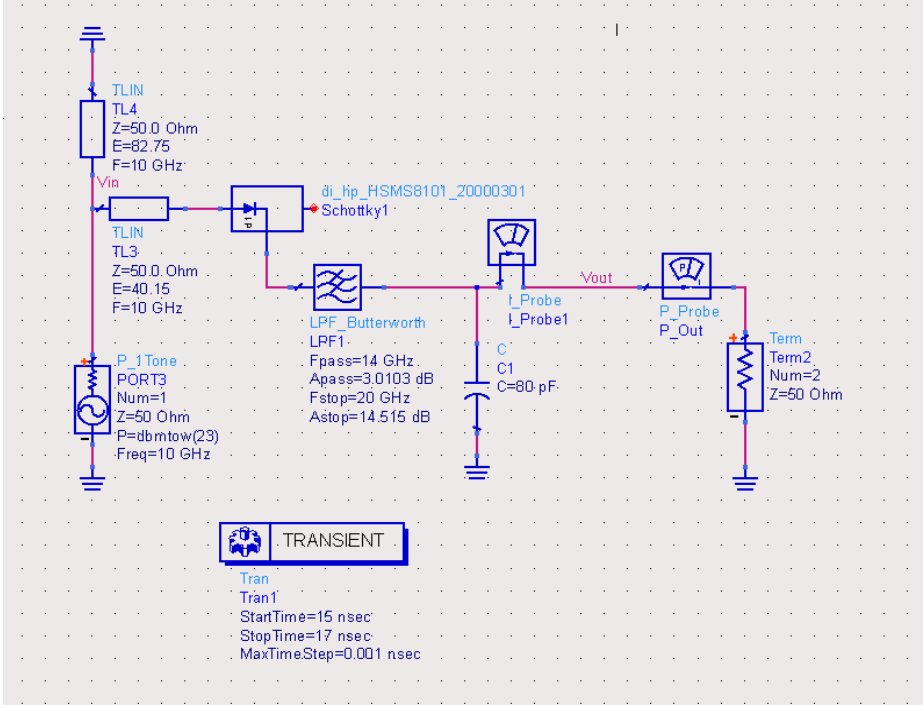

Figure 23. ADS model of a half-wave rectenna design with a post-LPF.

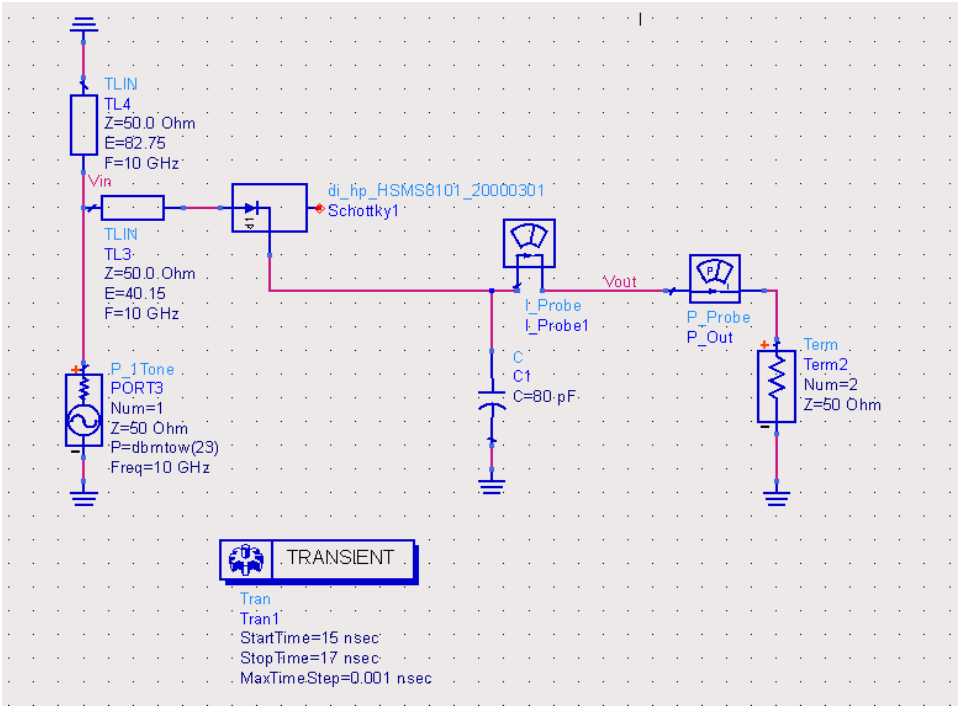

Figure 24. ADS model of a half-wave rectenna design without LPF.

All three designs utilize a capacitor after the diode. In reference [21], Vitale noted that the output voltage when the diode is not conducting is compensated for by the discharge of the filter capacitor. It can be expressed as

$$V_{out}(t) = V_P e^{-t/R_L C} \tag{3.4}$$

where V_P is the peak voltage of the rectified signal, R_L is the load resistance, and C is the filter capacitance. For $R_L C \gg T$, the exponential can be approximated by $\left(1 - T \big/ R_L C\right)$ and the minimum output voltage can be expressed as

$$V_{\min} = V_P \left(1 - \frac{T}{R_L C}\right) \tag{3.5}$$

where T is a full period of the input sinusoid. He also determined that the output dc voltage V_{out} is approximated by the average voltage between the peak voltage and the minimum output voltage

$$V_{out} = V_P \left(1 - \frac{1}{2 f R_L C}\right). \tag{3.6}$$

Figures 25 to 27 illustrate the output dc power produced by the three different circuit designs of a half-wave rectenna that were simulated in ADS.

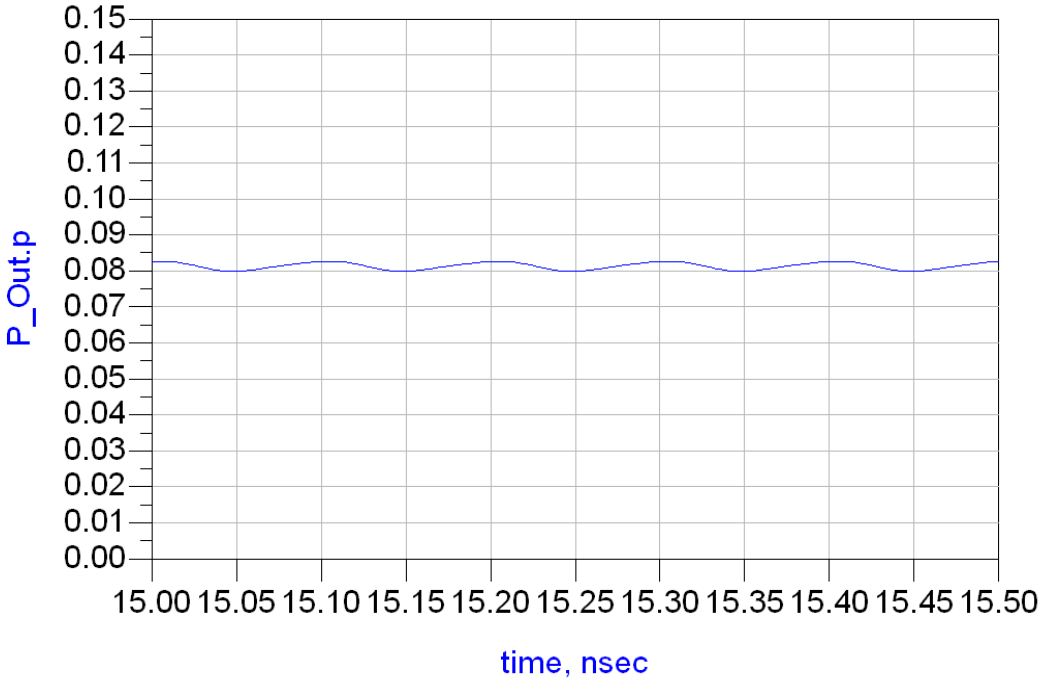

Figure 25. Output power (watts) versus time for a half-wave rectenna design with a pre-LPF.

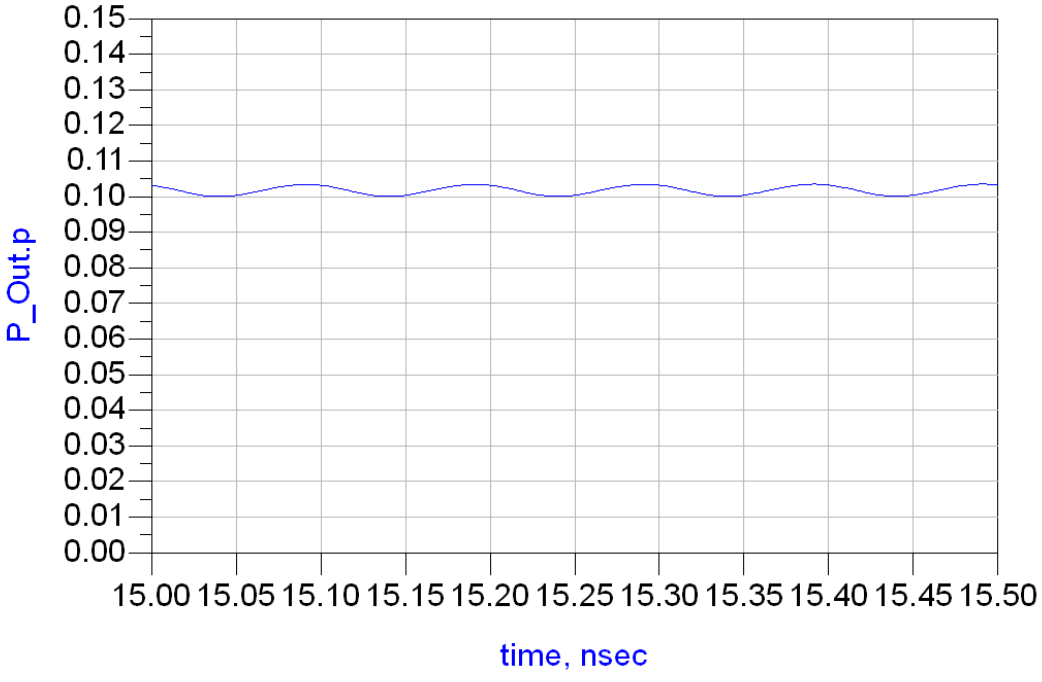

Figure 26. Output power (watts) versus time for a half-wave rectenna design with a post-LPF.

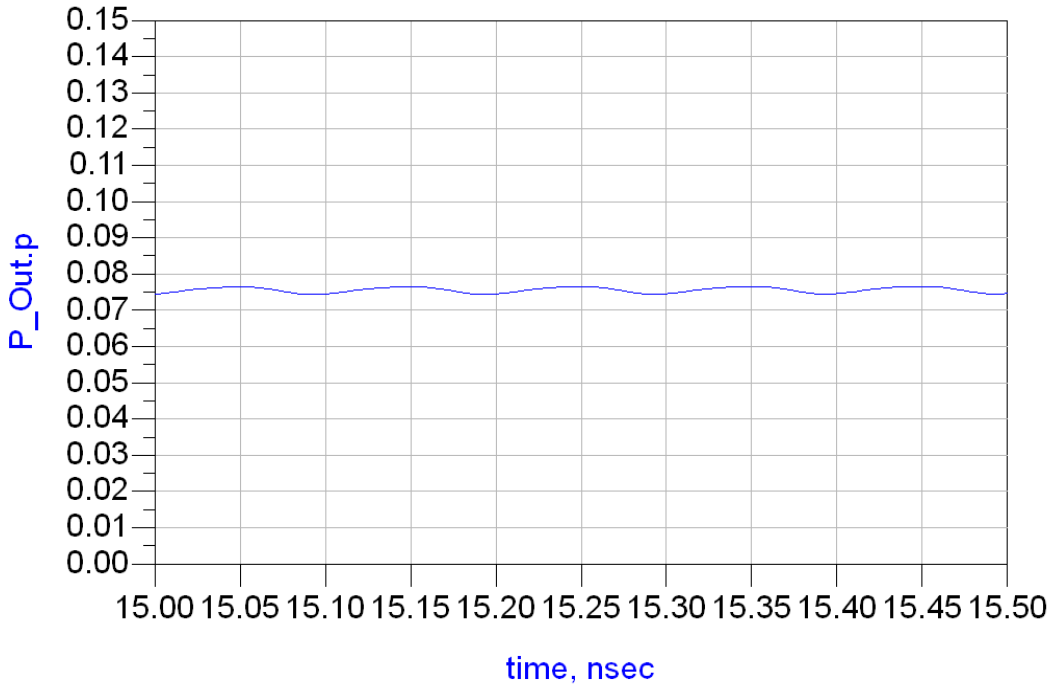

Figure 27. Output power (watts) versus time for a half-wave rectenna design without LPF.

In addition, rf-to-dc conversion efficiency η in percent is calculated as follows:

$$\eta = \frac{(V_{DC}/R_L)}{P_{RF}} \times 100 \tag{3.7}$$

where V_{DC} is dc output voltage, R_L is load resistance, and P_{RF} is microwave input power.

The simulated conversion efficiency of the half-wave rectennas is compared by using Equation (3.7). In Figure 28, the efficiency of the rectenna with post-LPF is higher than those for the pre-LPF and no LPF at their highest power levels. In comparison to the other designs shown in Figure 28, the design with post-LPF is able to convert wireless power to dc power with an efficiency of 54.3% at an input power of 150 mW, but the design with pre-LPF only achieves an efficiency of 45.8% at the same input power. In specific applications with input power smaller than 100 mW, we prefer the half-wave-rectification rectenna with pre-LPF to achieve a high conversion efficiency.

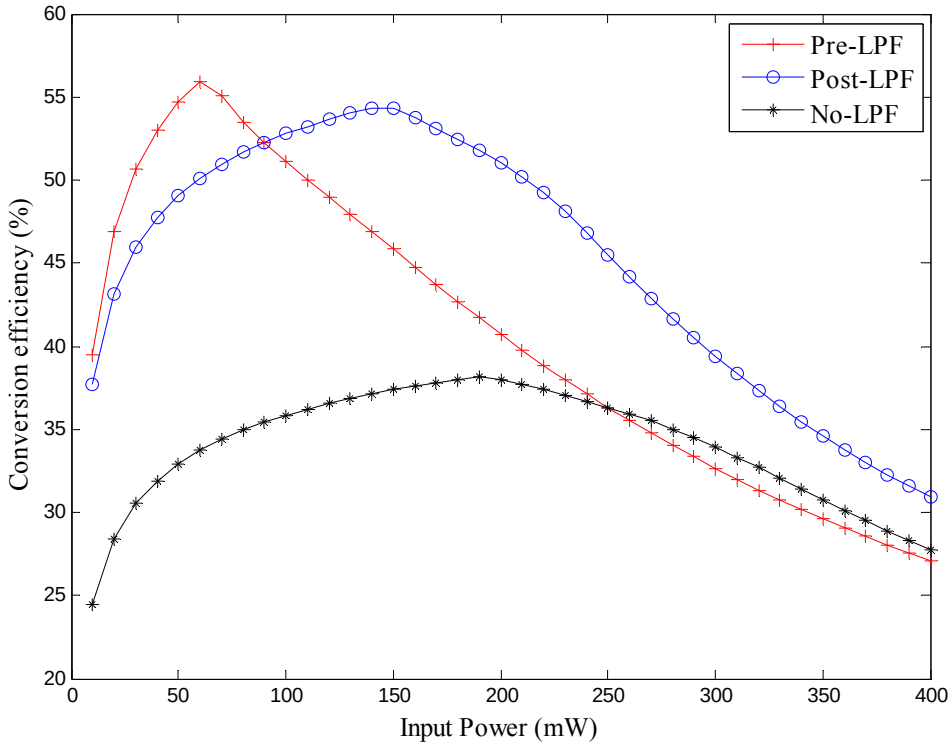

Figure 28. Conversion efficiency of simulated half-wave rectenna designs using ADS.

2. Hybrid Rectenna Design

In reference [18], Zbitou developed hybrid rectenna that achieved a conversion efficiency of 56% at 2.45 GHz. The block diagram of the hybrid rectenna is depicted in Figure 29. The function of the second diode is to provide a variable resistance in series with the load.

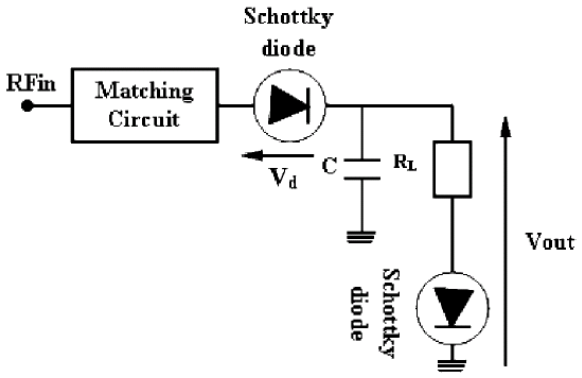

Figure 29. Block diagram of hybrid rectenna (From [18]).

Zbitou considered the dc equivalent circuit of the Schottky diode as a voltage source in series with the junction resistance R_j and that the output voltage of the rectifier can expressed as

$$V_{out} = V_{DC} \frac{R_L}{R_L + R_j} \tag{3.8}$$

where V_{DC} is the dc part of the diode junction voltage and R_j is the junction resistance of the Schottky diode. R_j is obtained by differentiating the diode voltage-current characteristics and is given by [18]

$$R_j = \frac{nKT}{q\left(I_s + I_b\right)} \tag{3.9}$$

where n is the diode ideality factor, K is Boltzmann's constant, q is the electronic charge, I_s is the diode saturation current, I_b is the external bias current, and T is the temperature of the diode in degrees Kelvin. For the HSMS2820 diode, R_j is approximately 1.7 MΩ at ambient temperature, which tends to decrease the value of V_{out}, as evident from Equation (3.8). According to Equation (3.8), an increase in voltage at low power levels can be obtained when a variable resistance, which fluctuates as R_j, is connected in series with R_L. The value of this variable resistor must change with the

rectified current. Thus, Zbitou employed a technique using the same Schottky diode in series with R_L, in the dc portion of the rectifier, to act as a variable resistor (due to its current dependence in the junction resistance).

Three variations of the hybrid circuit are evaluated. Two incorporate an LPF (Figures 30 and 31). The third eliminates the LPF in order to minimize size and weight (Figure 32).

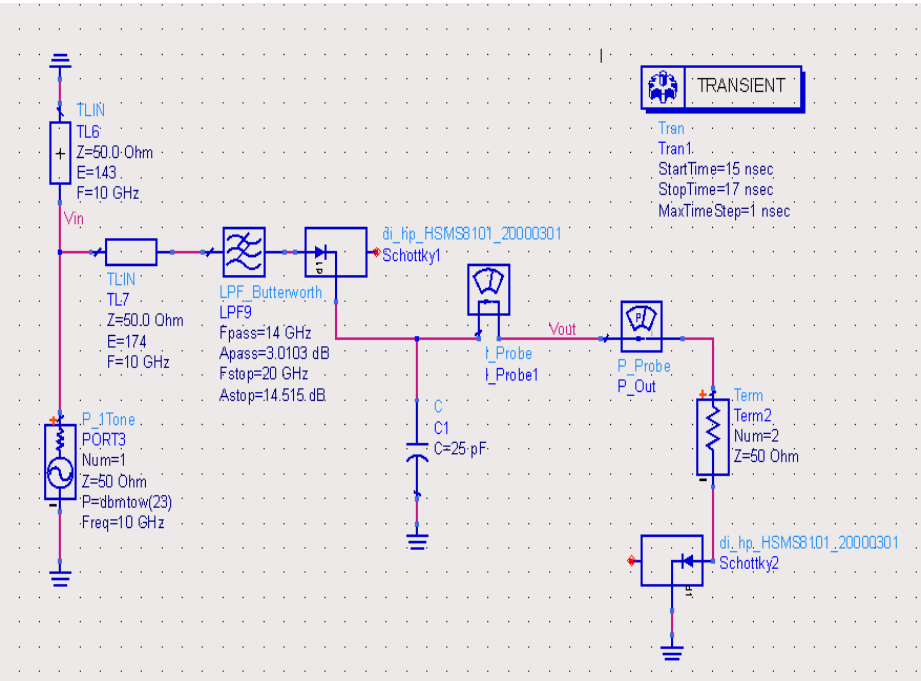

Figure 30. ADS model of a hybrid rectenna design with a pre-LPF.

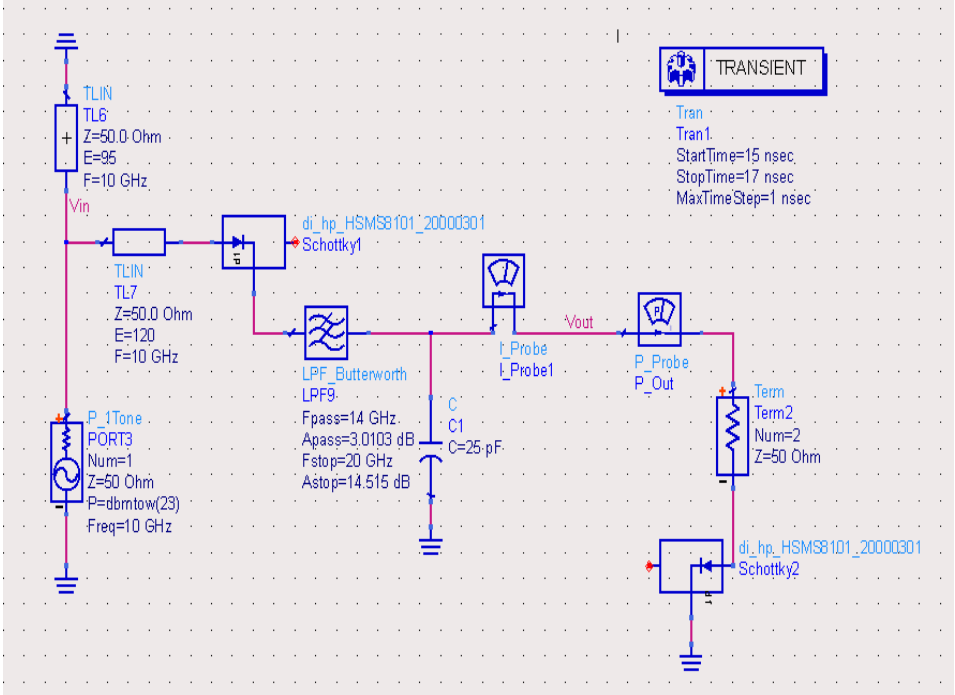

Figure 31. ADS model of a hybrid rectenna design with a post-LPF.

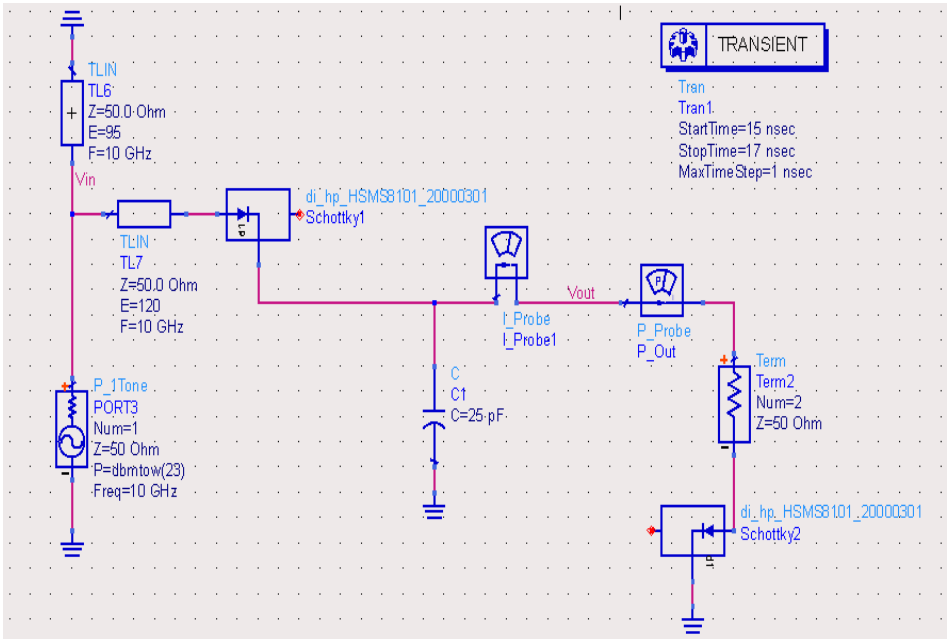

Figure 32. ADS model of a hybrid rectenna design without LPF.

Figures 33 to 35 illustrate the output dc power produced by three different circuit designs of hybrid rectenna that were simulated in ADS.

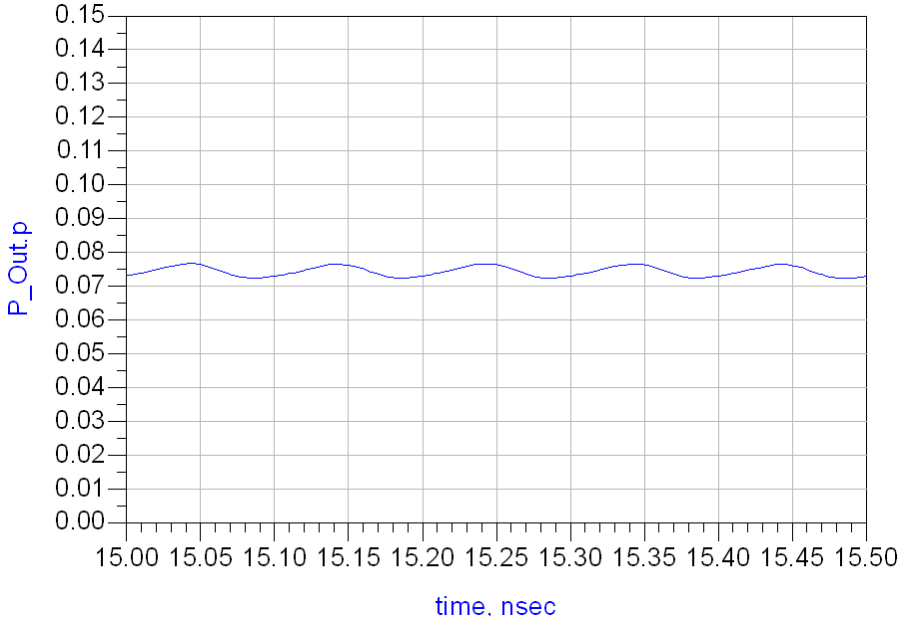

Figure 33. Output power (watts) versus time for a hybrid rectenna design with a pre-LPF.

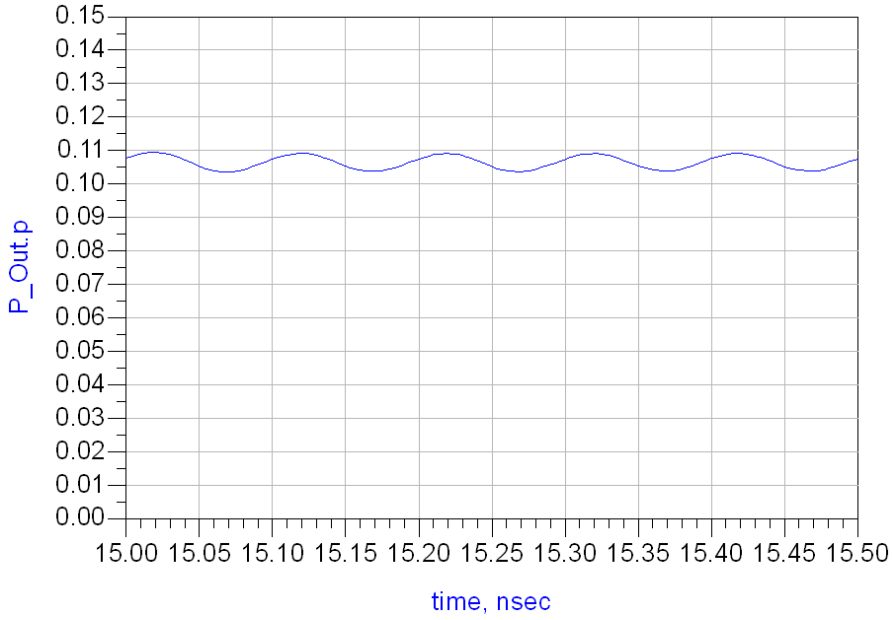

Figure 34. Output power (watts) versus time for a hybrid rectenna design with a post-LPF.

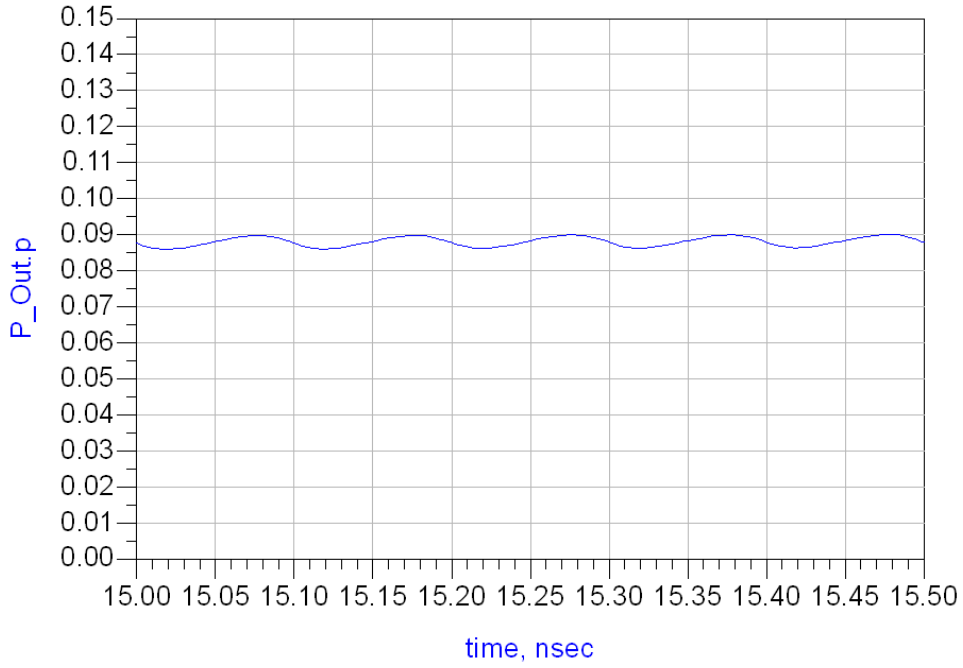

Figure 35. Output power (watts) versus time for a hybrid rectenna design without LPF.

The simulated conversion efficiency of the three hybrid rectenna designs are given in Figure 36. The efficiency of the hybrid rectenna with post-LPF is higher than those with pre-LPF and no-LPF. In comparison to the previous results given in Figure 36, the design with post-LPF is able to convert wireless power to dc power with an efficiency of 61.2% at an input power 140 mW, but the design with pre-LPF only achieves an efficiency of 45% at the same input power.

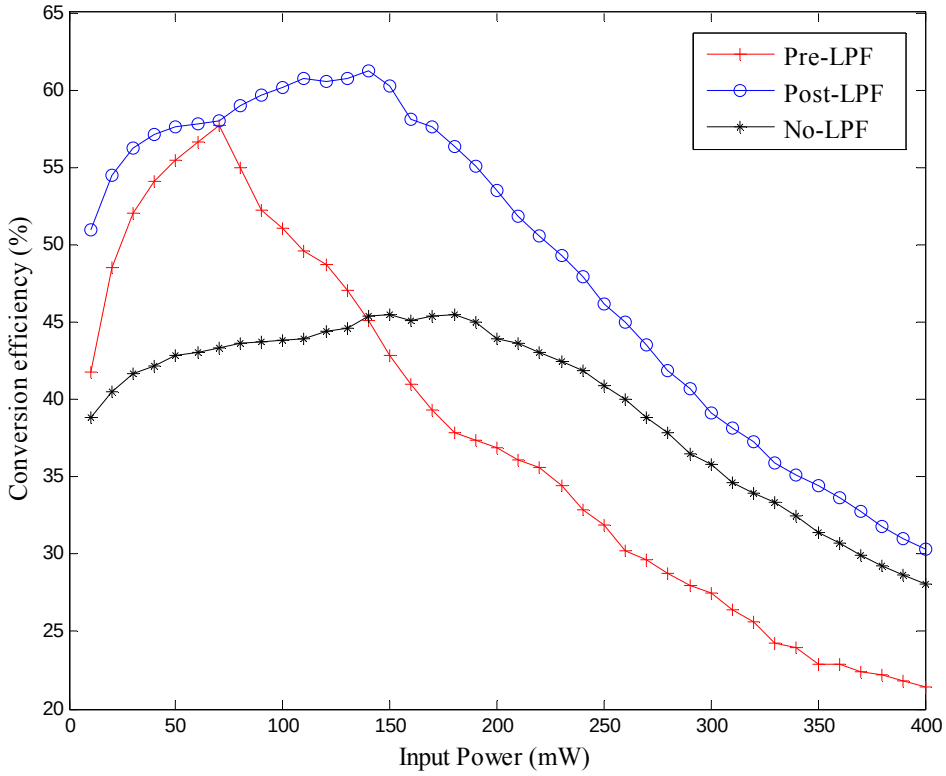

Figure 36. Simulated conversion efficiency of the hybrid rectennas.

3. Full-Wave Rectenna Design

In reference [23], Chiou presented a full-wave rectenna circuit in order to obtain a more stable dc output voltage than that a half-wave rectenna of the same chip area. He demonstrated the feasibility of a rectenna with an efficiency of 53% at an incident radiation power density of 30 W/cm^2 and frequency of 35 GHz. The rectenna comprised a power-receiving linear tapered slot antenna (LTSA), a slot line (SL) to finite-width ground coplanar waveguide (FGCPW) transition, a bandpass filter (BPF), a full-wave rectifier for rf-to-dc conversion, a dc bypass capacitor, and a resistive load. The fabricated rectenna with off-chip lumped elements is depicted on Figure 37.

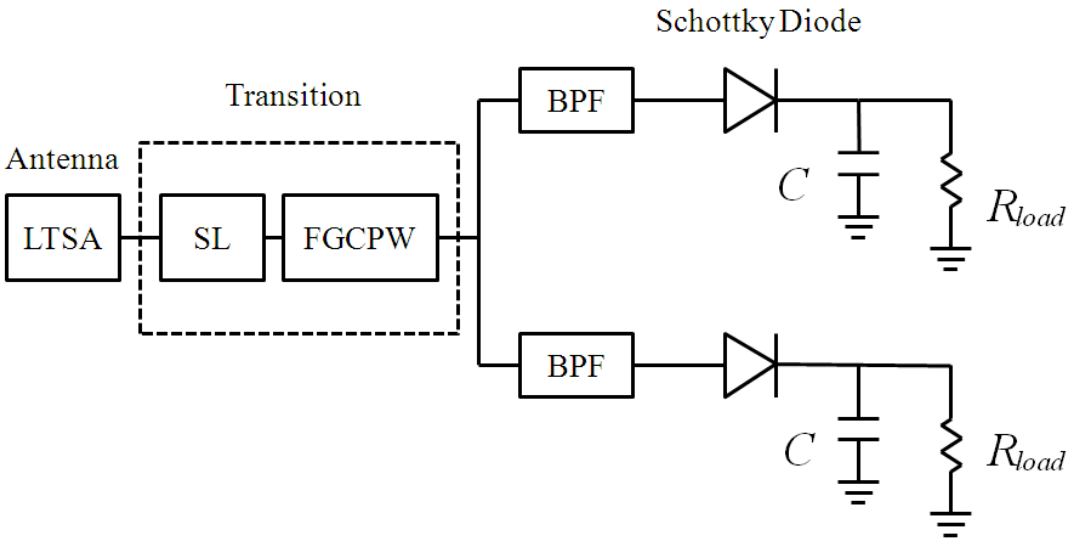

Figure 37. Block diagram of full-wave rectifier (From [23]).

As in the previous circuit, in order to minimize the size and weight, elimination of the low-pass filter is considered. For our design, a microstrip-fed dipole will be used in place of the LTSA. It is a lighter-weight structure that can be tuned to 10 GHz, whereas the LTSA is a broadband structure. Figures 38 to 40 illustrate the three different circuit designs for a full-wave rectenna.

In the ADS model, the antenna is represented by a power source with 50 Ω impedance. The diodes are matched using shorted stubs placed the required distance from the diodes. The stub location and stub length were computed using the standard stub-matching procedure [24]. Then the ADS optimization program was used to adjust the line lengths so the output power was a maximum.

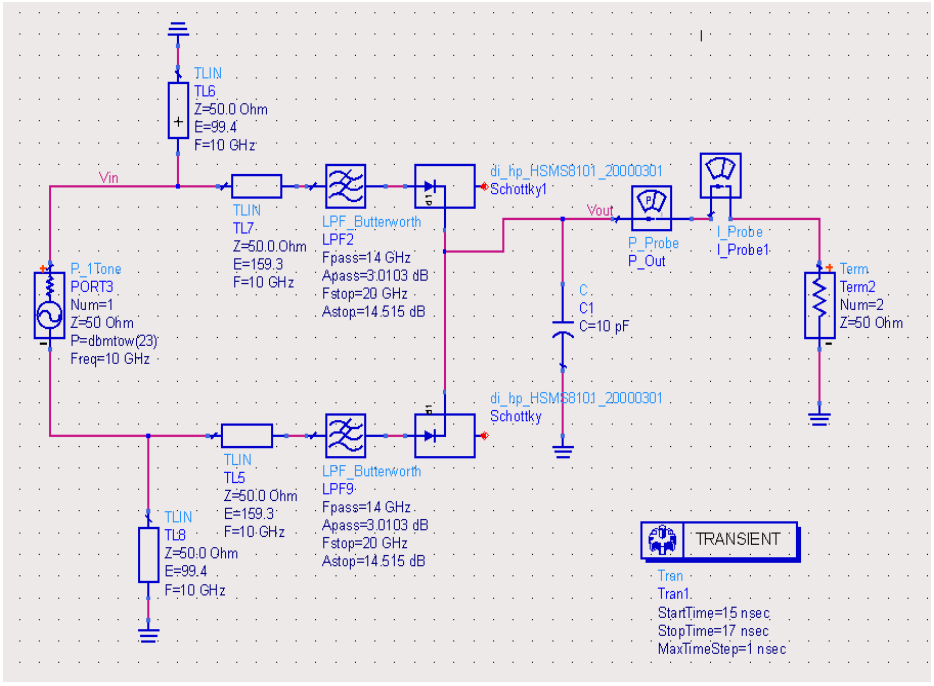

Figure 38. ADS model of a full-wave rectenna design with a pre-LPF.

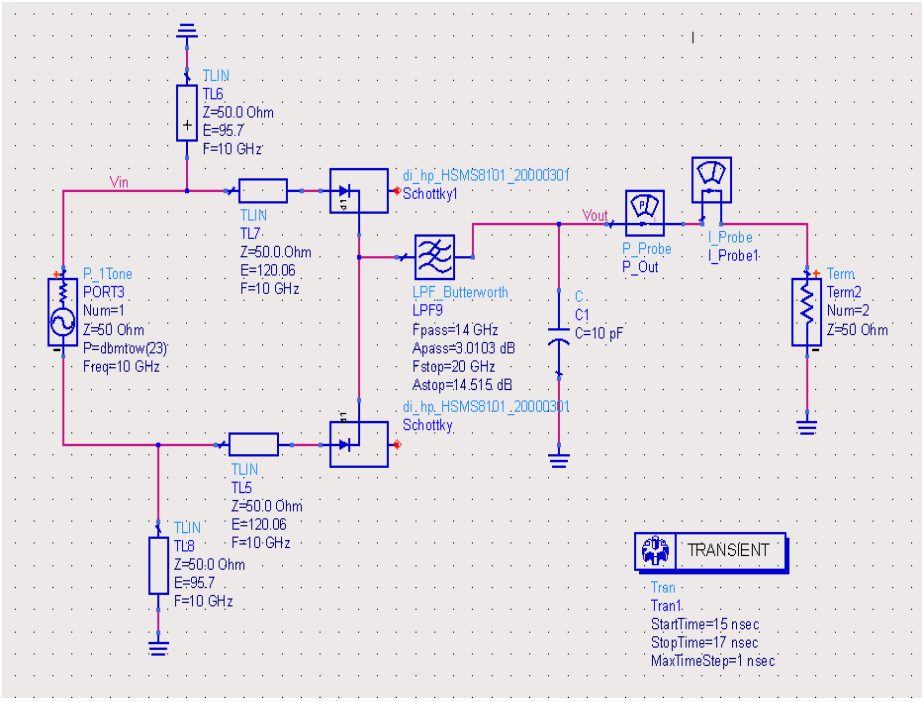

Figure 39. ADS model of a full-wave rectenna design with a post-LPF.

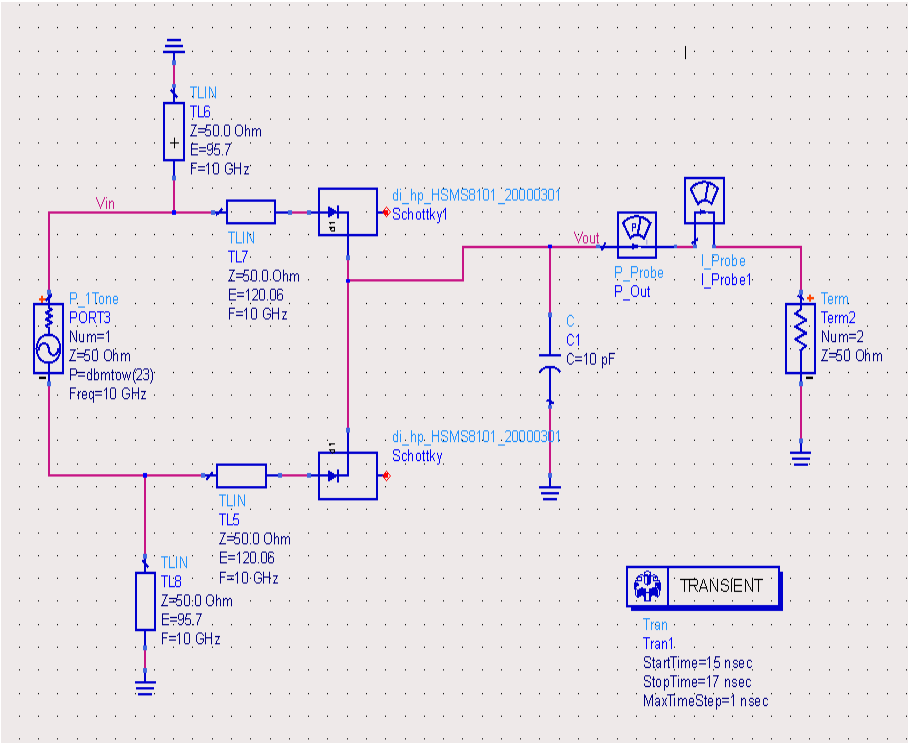

Figure 40. ADS model of a full-wave rectenna design without LPF.

The same method used to obtain Equation (3.4) was applied to the full-wave rectenna for the discharge time. The derivative expression for the output voltage V_{out} becomes

$$V_{out} = V_P \left(1 - \frac{1}{4 f R_L C} \right). \tag{3.10}$$

Equation (3.6) and (3.10) show that the average output voltage of the full-wave rectenna can approach double the half-wave rectenna.

Figures 41 to 43 illustrate the output dc power produced by three circuit designs of the full-wave rectenna that were simulated in ADS.

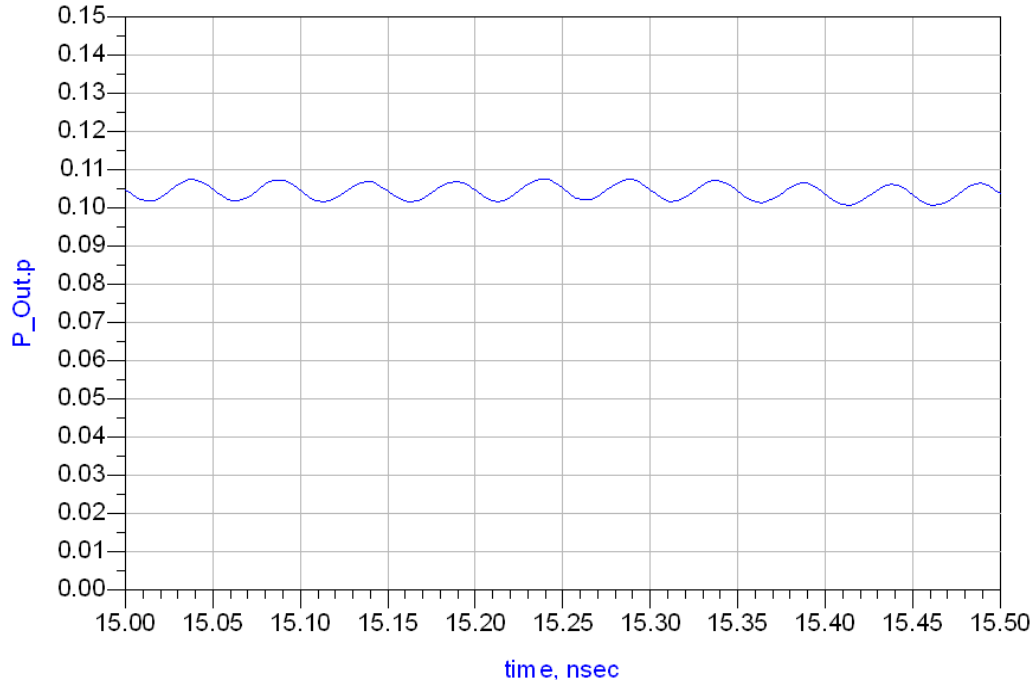

Figure 41. Output power (watts) versus time for a full-wave rectenna design with a pre-LPF.

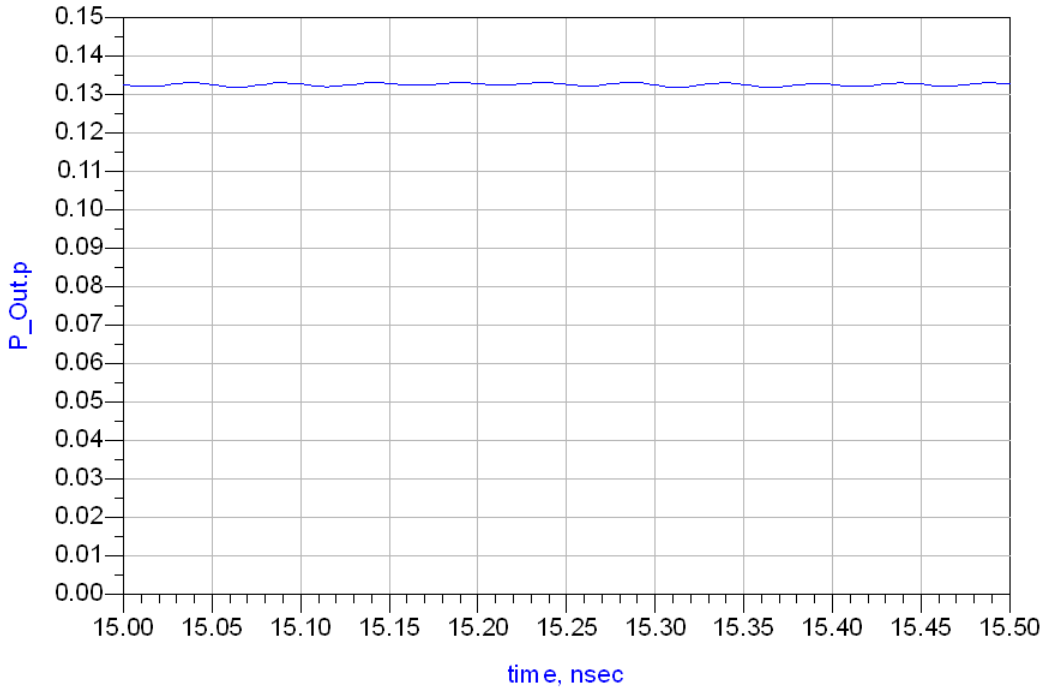

Figure 42. Output power (watts) versus time for a full-wave rectenna design with a post-LPF.

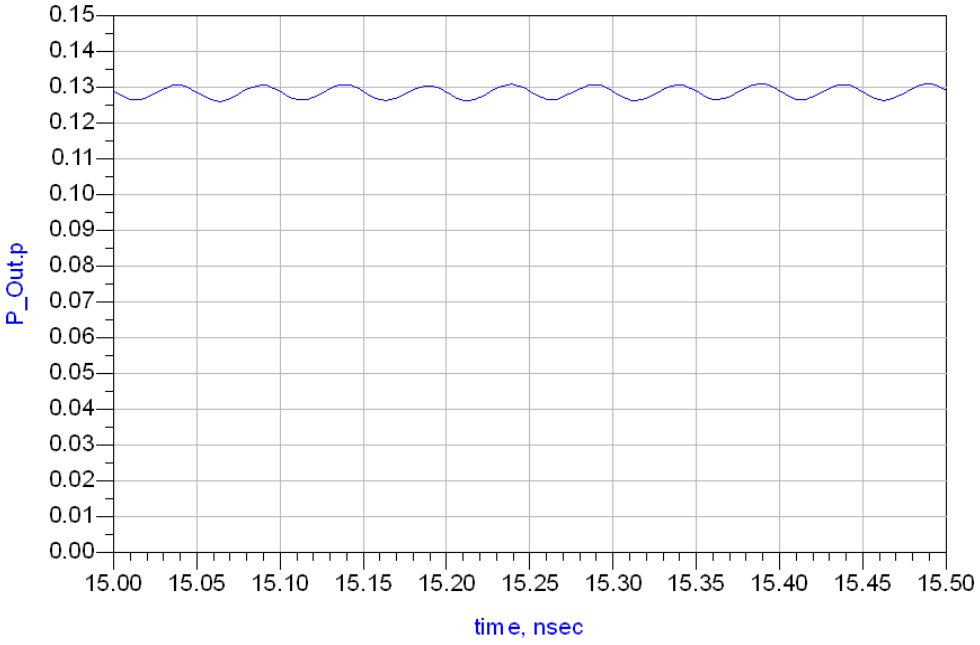

Figure 43. Output power (watts) versus time for a full-wave rectenna design without LPF.

The simulated conversion efficiencies of the three full-wave rectenna designs are given in Figure 44. In Figure 44, the efficiency of the full-wave rectenna with post-LPF and no-LPF are the same and higher than that with pre-LPF. In comparison with the previous results in Figure 33, the design with no LPF is able to convert wireless power to dc power with an efficiency of 68.1% at an input power of 170 mW, but the design with pre-LPF only achieves an efficiency of 56.6% at the same input power.

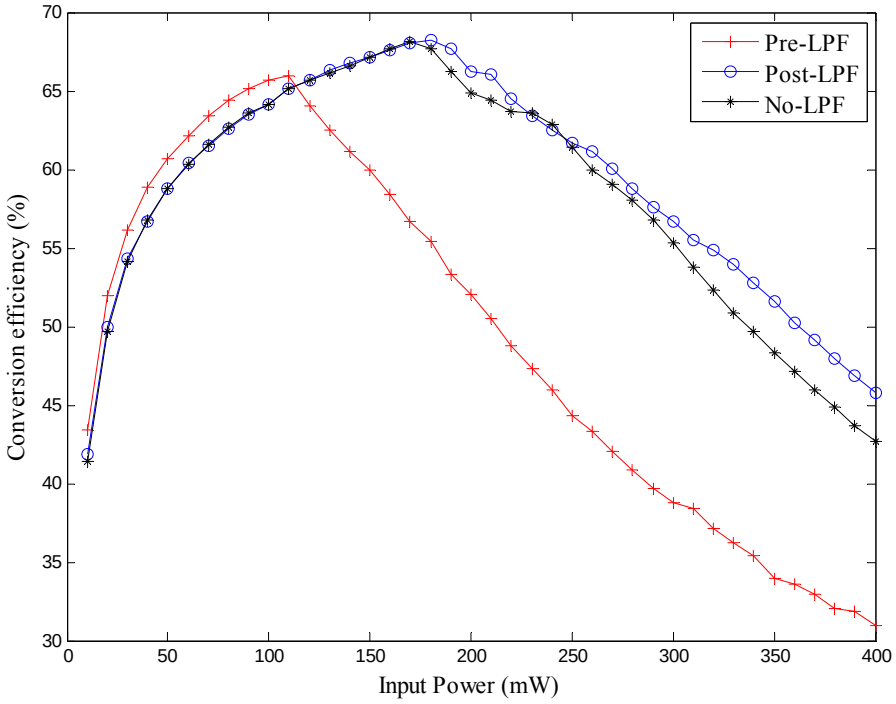

Figure 44. Simulated conversion efficiency of full-wave rectenna using ADS with different LPF.

4. Conclusion

Comparing previous results, we observed that the conversion efficiency of the hybrid rectenna is higher than that of the half-wave rectenna, which confirms the research conducted by Zbitou [18]. In order to hover the MAV, the full-wave rectenna with no LPF is selected to achieve the highest efficiency (68.1%) of the various rectenna designs and obtain more stable dc power than with the half-wave rectenna. The full-wave rectenna with no LPF also provides a lightweight design for the MAV application.

C. EVALUATION OF HARMONIC FREQUENCIES

In reference [12], Hagerty reported that reflected harmonic energy from the input or output side of the diode can alter the voltage across the diode. The diode also begins to bias itself as it produces more dc current, thus moving the dc operating point of the I-V curve in a nonlinear fashion. The diode's harmonic frequency components can possibly be radiated by the antenna, causing interference with other systems. Based on the

properties of the diode at microwave frequencies, we simulate and analyze the radiated harmonics and dc power of different rectenna designs for an input microwave power at 10 GHz. This can be accomplished using the harmonic balance (HB), nonlinear-circuit analysis module of the ADS software.

1. Harmonic Balance Analysis for the Half-Wave Rectenna

Figures 45 to 46 illustrate the harmonic balance of a half-wave rectenna with pre-LPF. The power in the fundamental frequency, the second through fourth harmonics, and the dc power are shown at four locations in the circuit (at the probes highlighted in the figure). These are the same cases shown in Figures 25 through 27, and their voltage and power curves were shown previously.

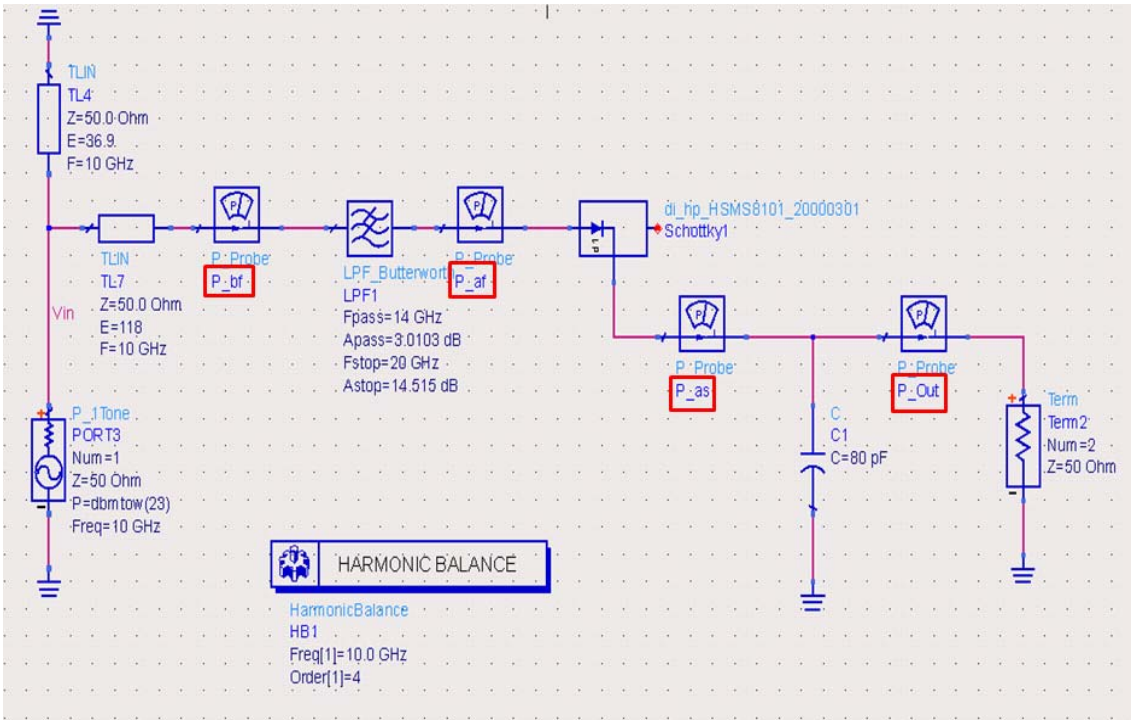

Figure 45. Harmonic balance of configuration in half-wave rectenna with pre-LPF.

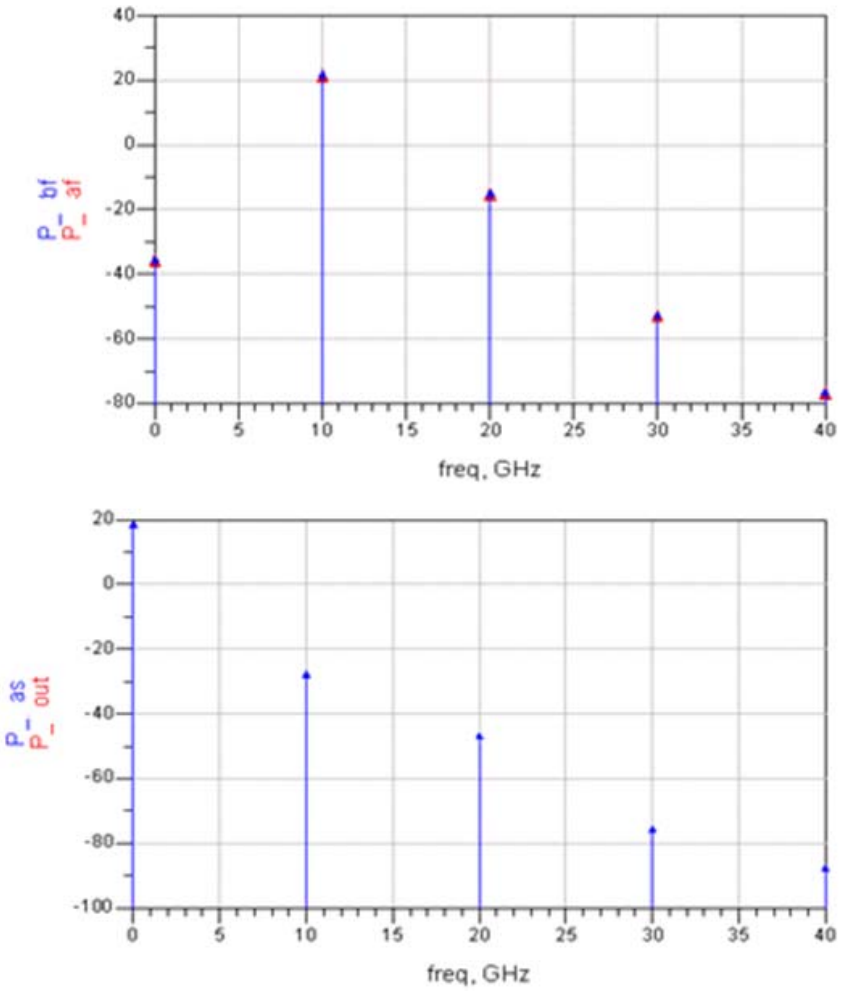

Figure 46. Simulated input power, reradiated harmonic powers, and dc power (in dBm) versus frequency for the half-wave rectenna with pre-LPF.

The simulation results for the power in the harmonics at different nodes in the circuit are given in Table 3. Pbf is the power before the low-pass filter, Paf is the power after the low-pass filter, Pbs is the power after the Schottky diode, and Pout is the power at the output of the circuit. The unit is dBm. Examination of the results in Table 3 shows that the harmonics reflected by diode are eliminated by the low-pass filter. The 80 pF capacitor performs as a dc pass filter to eliminate any leakage of 10 GHz signal. We confirm the above simulation by comparing the dc power of the harmonic balance simulation with that of the transient simulation, and found that the two results agree.

Harmonic balance analysis data for fundamental frequency of 10 GHz (Pin=23 dBm)					
Node	0 Hz	10 GHz	20 GHz	30 GHz	40 GHz
Pbf (dBm)	-34.302	22.887	-13.708	-51.469	-75.358
Paf (dBm)	-34.302	22.887	-13.708	-51.469	-75.358
Pas (dBm)	19.634	-26.804	-45.813	-74.595	-86.818
Pout (dBm)	19.634	-26.804	-45.813	-74.595	-86.818

Table 3. Harmonic balance data for the half-wave rectenna with pre-LPF.

Figures 47 to 48 illustrate the harmonic balance model of the half-wave rectenna with post-LPF and the power probes at different nodes in the circuit.

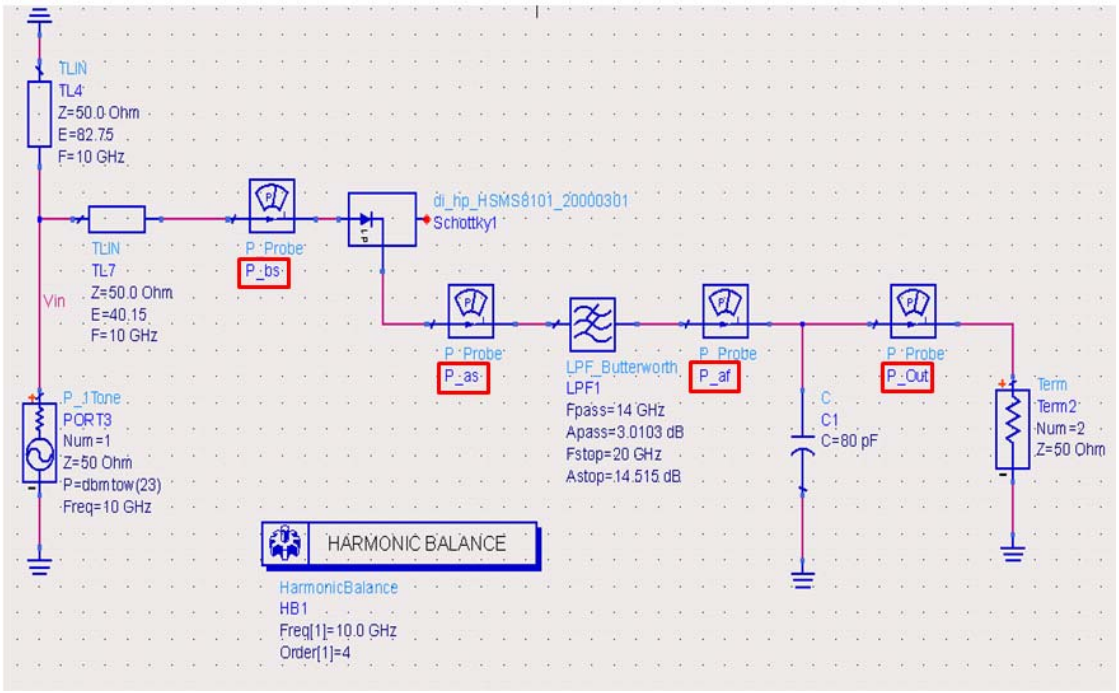

Figure 47. Harmonic balance simulation configuration for the half-wave rectenna with post-LPF.

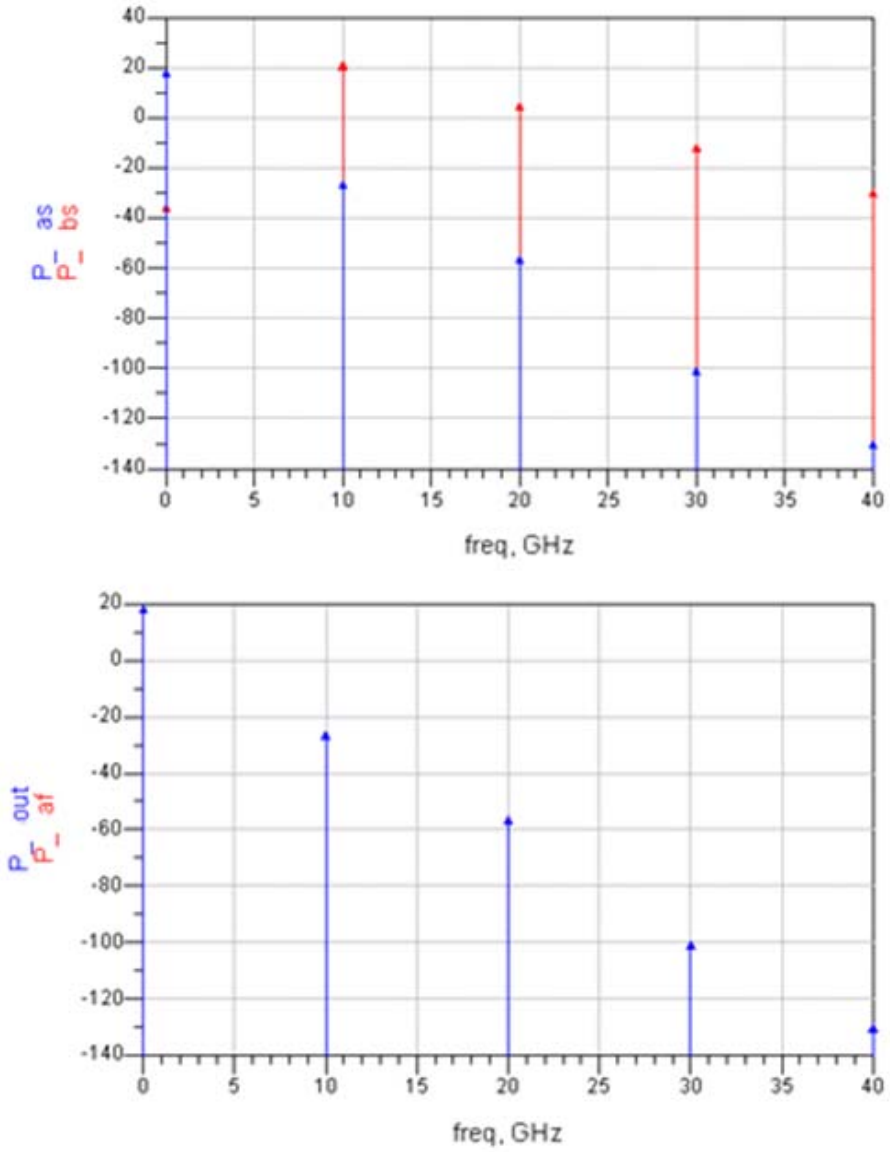

Figure 48. Simulated input power, reradiated harmonic powers, and dc power (in dBm) versus frequency for the half-wave rectenna with post-LPF.

Table 4 is a list of the power in the harmonics at different nodes in the circuit shown in Figure 45. Pbs is the power before the Schottky diode, Pas is the power after the diode, Paf is the power after the low-pass filter, and Pout is the power at the output. A comparison of the results in Table 4 shows that the highest harmonic reflected by diode and reradiated toward antenna was about 6 dBm at 20 GHz; the third and forth harmonics were negligible for this design. The low-pass filter and 80 pF capacitor reduced the

harmonics produced by diode and produced stable dc power. We confirmed the above simulation by comparing the dc power of harmonic balance simulation with the transient simulation, and the results agree.

Harmonic balance analysis data for fundamental frequency of 10 GHz (Pin=23 dBm)					
Node	0 Hz	10 GHz	20 GHz	30 GHz	40 GHz
Pbs (dBm)	-34.555	22.342	5.914	-10.423	-28.589
Pas (dBm)	19.381	-25.154	-55.312	-99.631	-129.019
Paf (dBm)	19.381	-25.154	-55.312	-99.631	-129.019
Pout (dBm)	19.381	-25.154	-55.312	-99.631	-129.019

Table 4. Harmonic balance data for the half-wave rectenna with post-LPF.

Figures 49 to 50 illustrate the harmonic balance of the half-wave rectenna without LPF and the power in the fundamental frequency, harmonics, and dc at different nodes in the circuit.

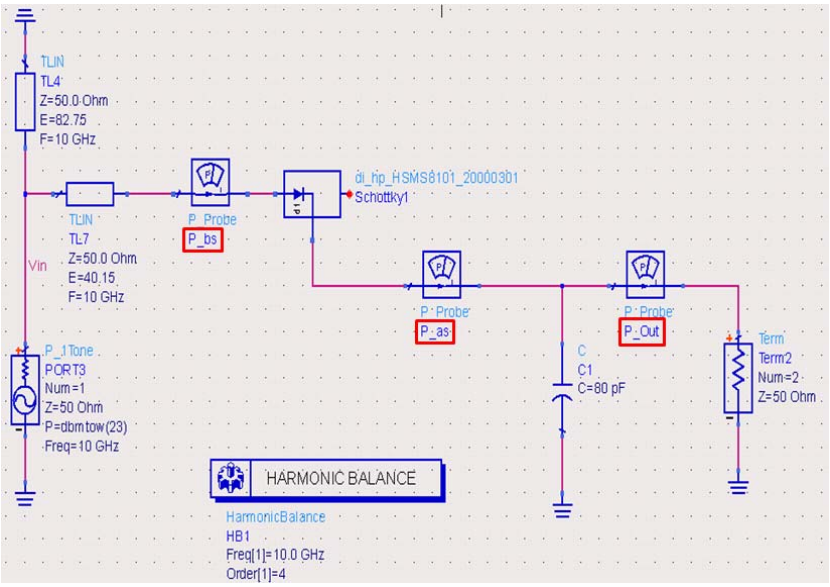

Figure 49. Harmonic balance simulation configuration for the half-wave rectenna without LPF.

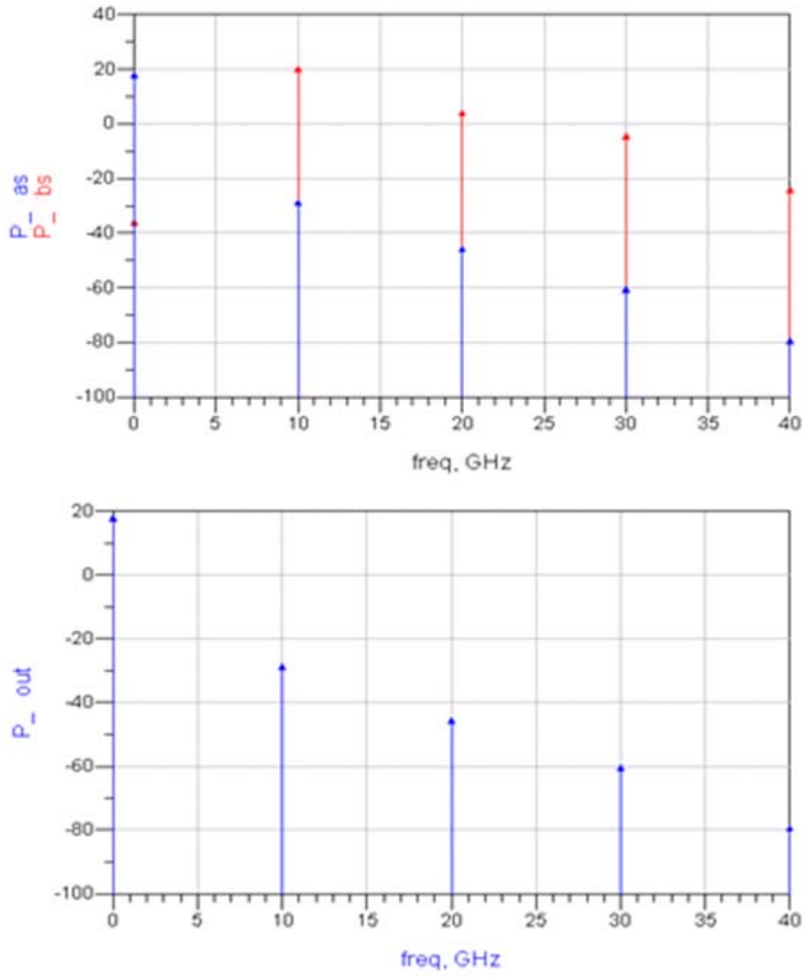

Figure 50. Simulated input power, reradiated harmonic power, and dc power (in dBm) versus frequency for the half-wave rectenna without LPF.

Table 5 is a list of the power in the harmonics at different nodes in the circuit. Pbs is the power before the Schottky diode, Pas is the power after the diode, and Pout is the power at the output. In comparison to the results in Table 5, the highest harmonic reflected by diode and reradiated toward antenna was about 5 dBm at 20 GHz, but the third and fourth harmonics were small and acceptable for our design. The 80 pF capacitor reduced the harmonics produced by diode and produced stable dc power. We confirmed the above simulation by comparing the dc power of the harmonic balance with the transient simulation, and the results agree.

Harmonic balance analysis data for fundamental frequency of 10 GHz (Pin=23 dBm)					
Node	0 Hz	10 GHz	20 GHz	30 GHz	40 GHz
Pbs (dBm)	-35.061	21.128	4.968	-3.422	-22.986
Pas (dBm)	18.876	-27.815	-44.730	-59.540	-78.615
Pout (dBm)	18.876	-27.815	-44.730	-59.540	-78.615

Table 5. Harmonic-balance data for the half-wave rectenna without LPF.

2. Harmonic Balance Analysis for the Full-Wave Rectenna

Figures 51 to 52 illustrate the harmonic balance of a full-wave rectenna with pre-LPF and the power in the fundamental frequency, harmonics, and dc at different nodes in the circuit.

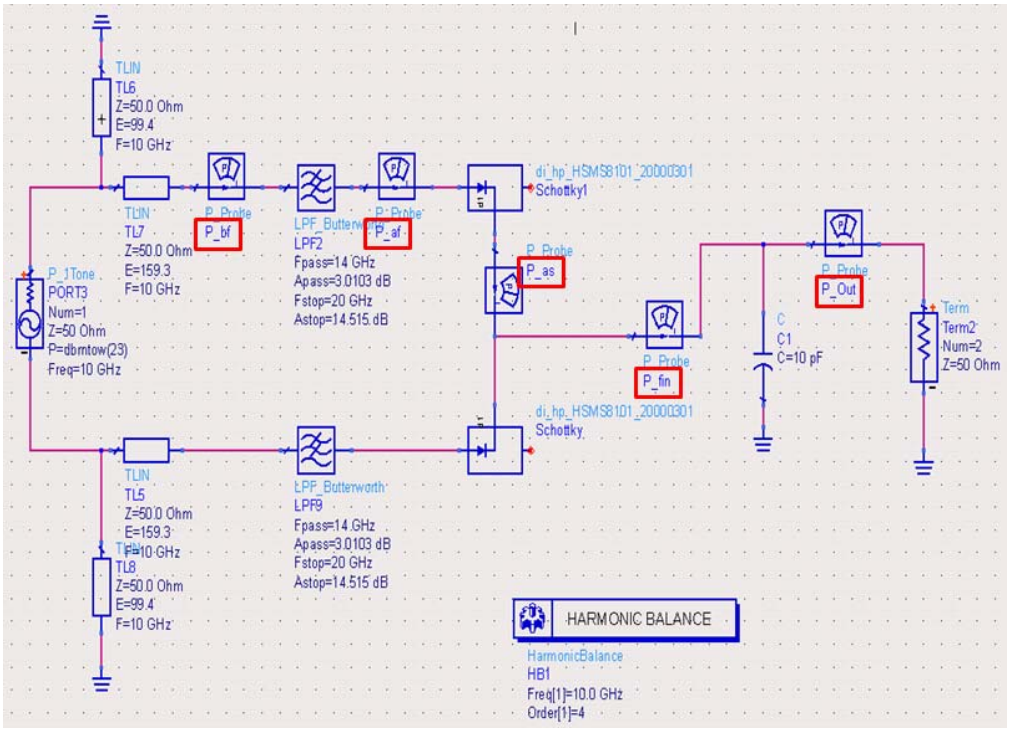

Figure 51. Harmonic-balance simulation configuration for the full-wave rectenna with pre-LPF.

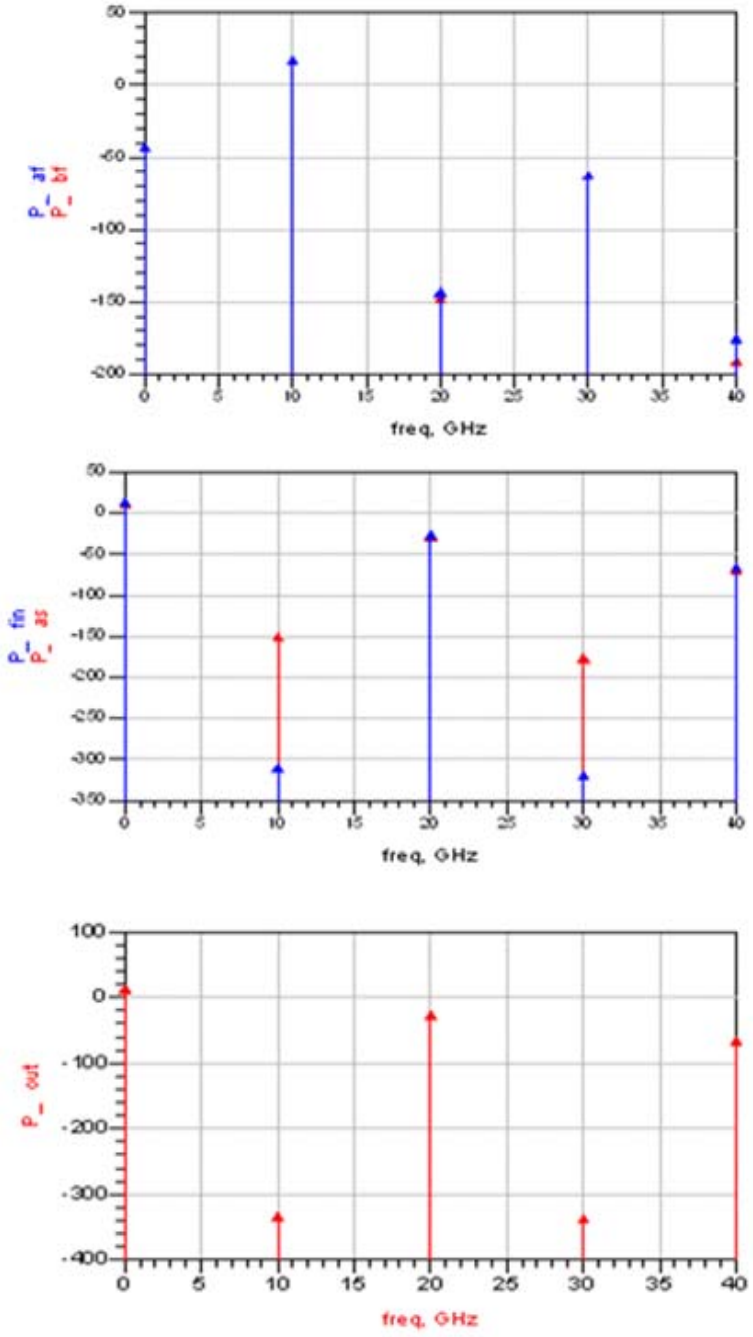

Figure 52. Simulated input power, reradiated harmonic power, and dc power (in dBm) versus frequency in full-wave rectenna with pre-LPF.

The power in the harmonics at different nodes in circuit is given in Table 6. Pbf is the power before the lowpass filter, Paf is the power after LPF, Pas is the power after the Schottky diode, Pfin is the combined power after the two diodes and Pout is the power at

the output of the circuit. The harmonics reflected by diode were eliminated by the low-pass filter. The full-wave architecture essentially eliminates the frequencies of 10 GHz and 30 GHz at the output. The 10 pF capacitor performs as a dc pass filter to eliminate the frequencies 20 GHz and 40 GHz. These simulation results were compared to the transient simulation, and the results agree.

Harmonic balance analysis data for fundamental frequency of 10 GHz (Pin=23 dBm)					
Node	0 Hz	10 GHz	20 GHz	30 GHz	40 GHz
Pbf (dBm)	-40.149	19.934	-145.177	-59.528	-188.391
Paf (dBm)	-40.149	19.934	-140.204	-59.528	-172.270
Pas (dBm)	16.797	-145.791	-23.490	-170.867	-63.062
Pfin (dBm)	19.808	-304.622	-20.480	-313.664	-60.051
Pout (dBm)	19.808	-326.624	-20.480	-330.596	-60.051

Table 6. Harmonic balance data for the full-wave rectenna with pre-LPF.

Figures 53 to 54 illustrate the harmonic balance of the full-wave rectenna with post-LPF and the power in the fundamental, harmonics, and dc at different nodes in circuit.

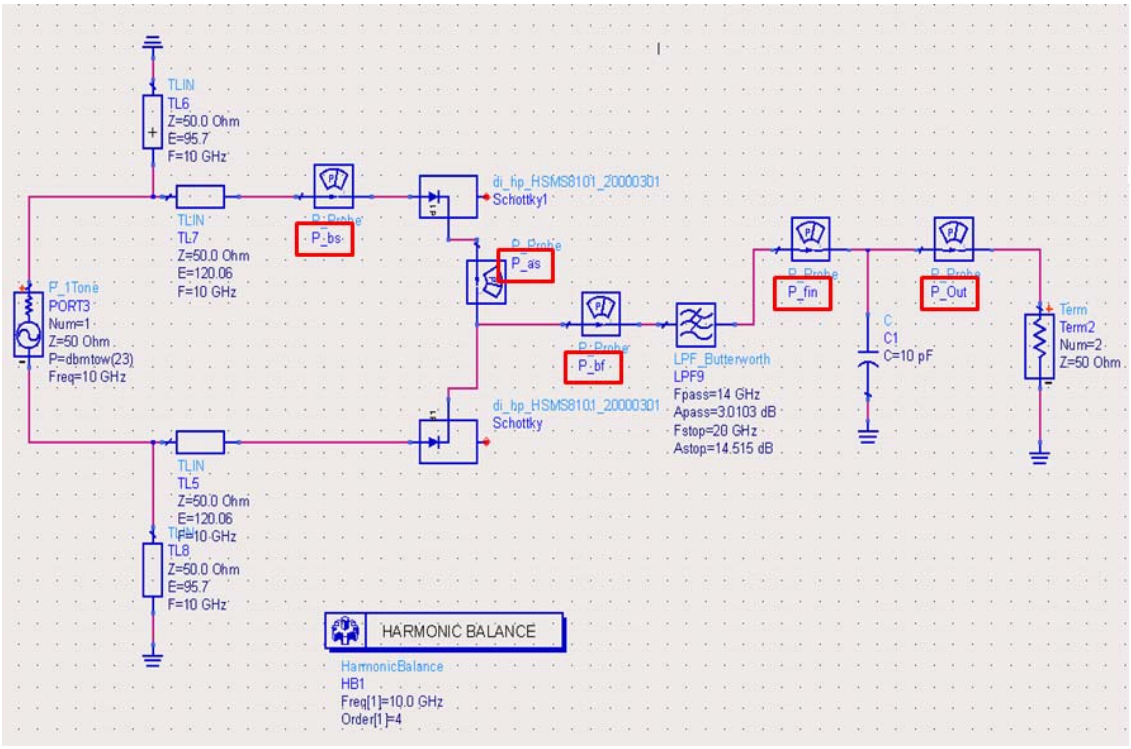

Figure 53. Harmonic-balance simulation configuration for the full-wave rectenna with post-LPF.

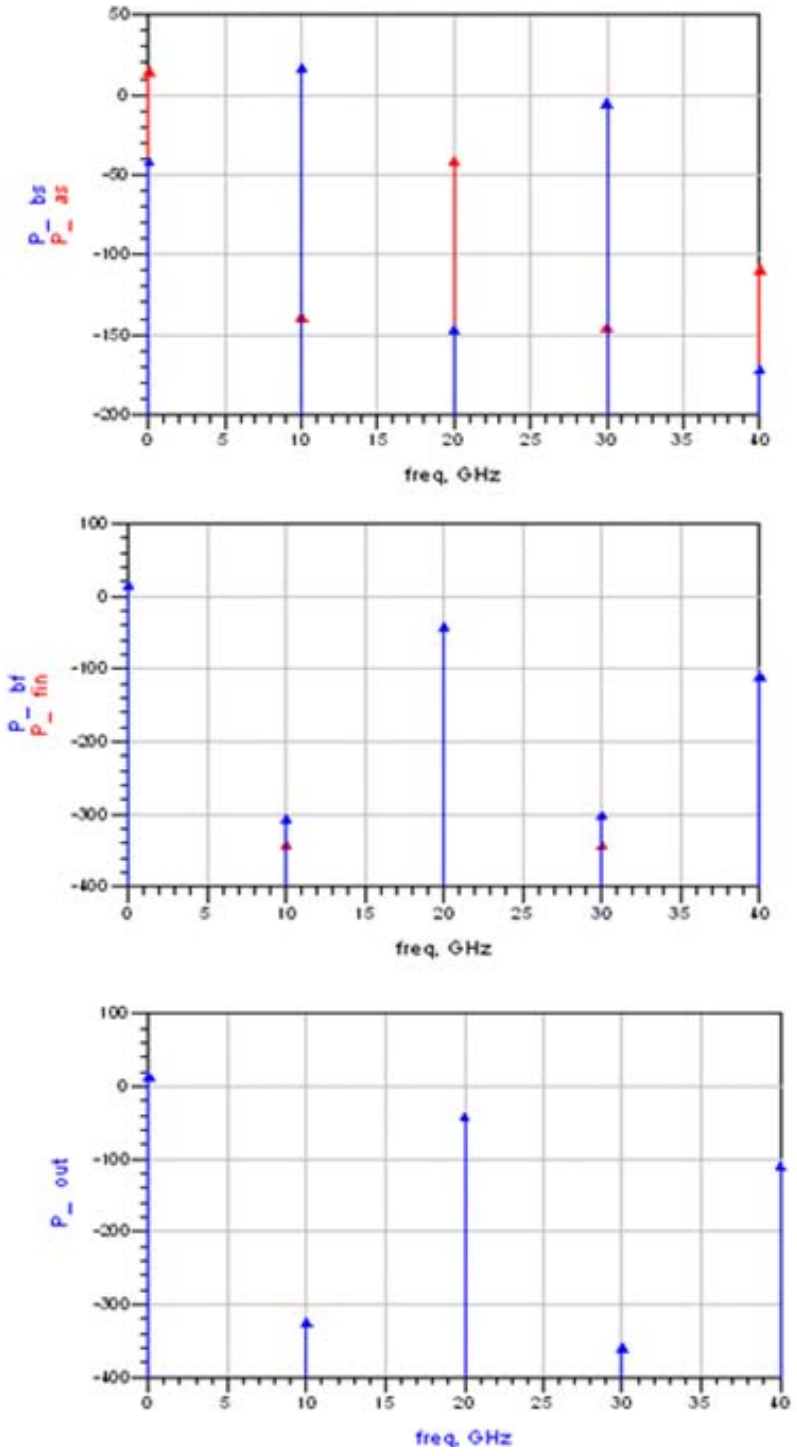

Figure 54. Simulated input power, reradiated harmonic power, and dc power (in dBm) versus frequency for the full-wave rectenna with post-LPF.

The power in the harmonics at different nodes in the circuit is given in Table 7, where Pbs is the power before the Schottky diode, Pas is the power after the diode, Pbf is the combined power of the two diodes and measured before the LPF, Pfin is the power after LPF, and Pout is the power at output. In comparison to the results of Table 7, the highest harmonic power reflected by the diode was -2 dBm at 30 GHz and there was negligible power at 10 GHz after the diode for the full-wave design. The full-wave architecture also eliminates the frequency of 30 GHz after the diode. The 10 pF capacitor performs as dc pass filter to eliminate the frequency of 20 GHz and 40 GHz. The simulation results were compared to the transient simulation, and the results agree with each other.

Harmonic balance analysis data for fundamental frequency of 10 GHz (Pin=23 dBm)					
Node	0 Hz	10 GHz	20 GHz	30 GHz	40 GHz
Pbs (dBm)	-39.176	19.839	-144.086	-2.080	-169.107
Pas (dBm)	17.770	-136.007	-38.575	-142.487	-106.223
Pbf (dBm)	20.781	-300.027	-35.564	-295.822	-103.212
Pfin (dBm)	20.781	-335.853	-35.564	-336.802	-103.212
Pout (dBm)	20.781	-317.919	-35.564	-352.415	-103.212

Table 7. Harmonic-balance data for the full-wave rectenna with post-LPF.

Figures 55 to 56 illustrate the harmonic balance of full-wave rectenna without LPF and the power in the fundamental frequency, harmonics, and dc at different nodes in circuit.

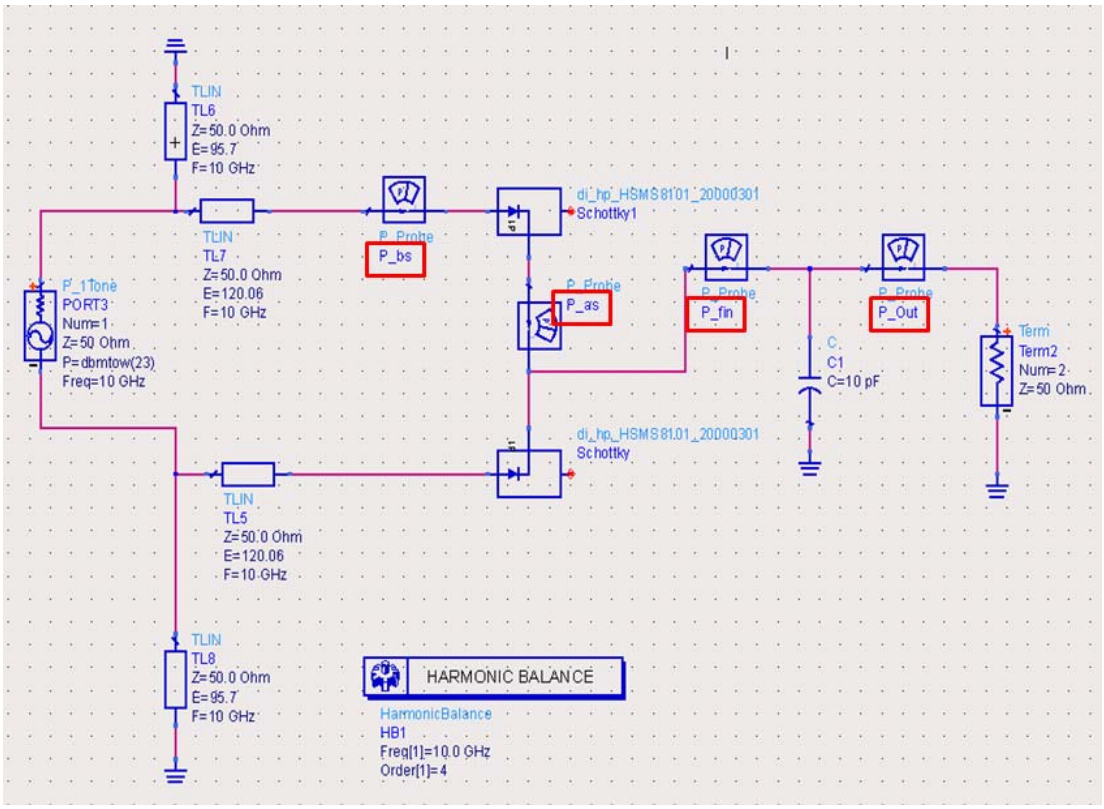

Figure 55. Harmonic-balance simulation configuration for the full-wave rectenna without LPF.

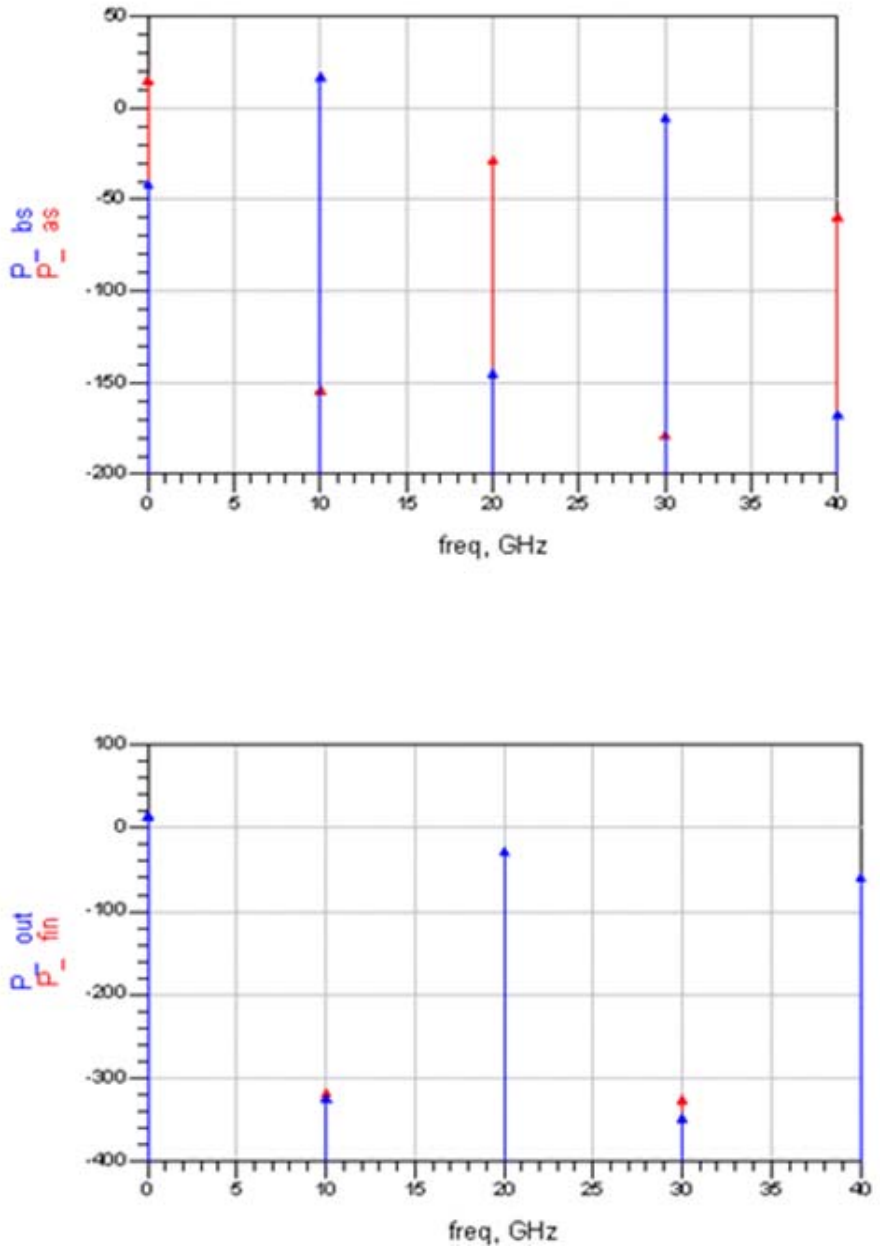

Figure 56. Simulated input power, reradiated harmonic power, and dc power (in dBm) versus frequency for the full-wave rectenna without LPF.

The simulation results for the power in the harmonics at different nodes in the circuit are given in Table 8. Pbs is the power before the Schottky diode, Pas is the power after the diode, Pfin is the combined power of the two diodes before the capacitor, and Pout is the power at the output. In comparison to the results in Table 8, the highest harmonic power generated by the diode was -3 dBm at 30 GHz and there was negligible

fundamental power at the output for the full-wave design. The 10 pF capacitor performs as a dc pass filter to eliminate the frequency of 20 GHz and 40 GHz. The simulation was compared to the transient simulation, and the results agree.

Harmonic balance analysis data for fundamental frequency of 10 GHz (Pin=23 dBm)					
Node	0 Hz	10 GHz	20 GHz	30 GHz	40 GHz
Pbs (dBm)	-39.277	19.824	-142.325	-3.127	-164.700
Pas (dBm)	17.670	-151.310	-26.147	-175.864	-56.734
Pfin (dBm)	20.680	-312.757	-23.137	-321.039	-53.723
Pout (dBm)	20.680	-318.761	-23.137	-343.172	-53.723

Table 8. Harmonic-balance data for the full-wave rectenna without LPF.

3. Findings

After comparing the three half-wave rectenna designs, we conclude that the half-wave rectenna with pre-LPF has better performance due to its limitation of harmonics back toward the antenna and slightly higher efficiency, albeit not much. The half-wave rectenna only converts at best half of the microwave energy, but the full-wave rectenna ideally converts all power, due to its architecture. We also observe that the full-wave rectenna without LFP converts more dc power than other designs and the full-wave rectifier design eliminates the fundamental frequency and third harmonic frequency at the output. In order to have a lightweight rectenna, the full-wave rectenna without LFP is selected. In the next chapter, we discuss the impedance of the full-wave rectification rectenna without LPF and optimize the impedance matching unit.

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IV. RECTENNA IMPLEMENTATION

A. RECTENNA IMPEDANCE DESIGN

Chapter III presented the simulation results for several different rectenna designs. We conclude that the full-wave rectenna without LPF meets the MAV requirement because of its higher conversion efficiency and constant output power. This chapter discusses the impedance matching of the rectenna and the circuit design and layout.

1. Full-Wave Rectenna Design

The full-wave rectenna design with the final optimized circuit and dielectric material is shown in Figure 57. The simulated output dc power, voltage, and current of full-wave rectenna are shown in Figure 58. The output power of the rectenna is 131 mW for an input power of 23 dBm, and the voltage oscillates from 2.55 V to 2.53 V.

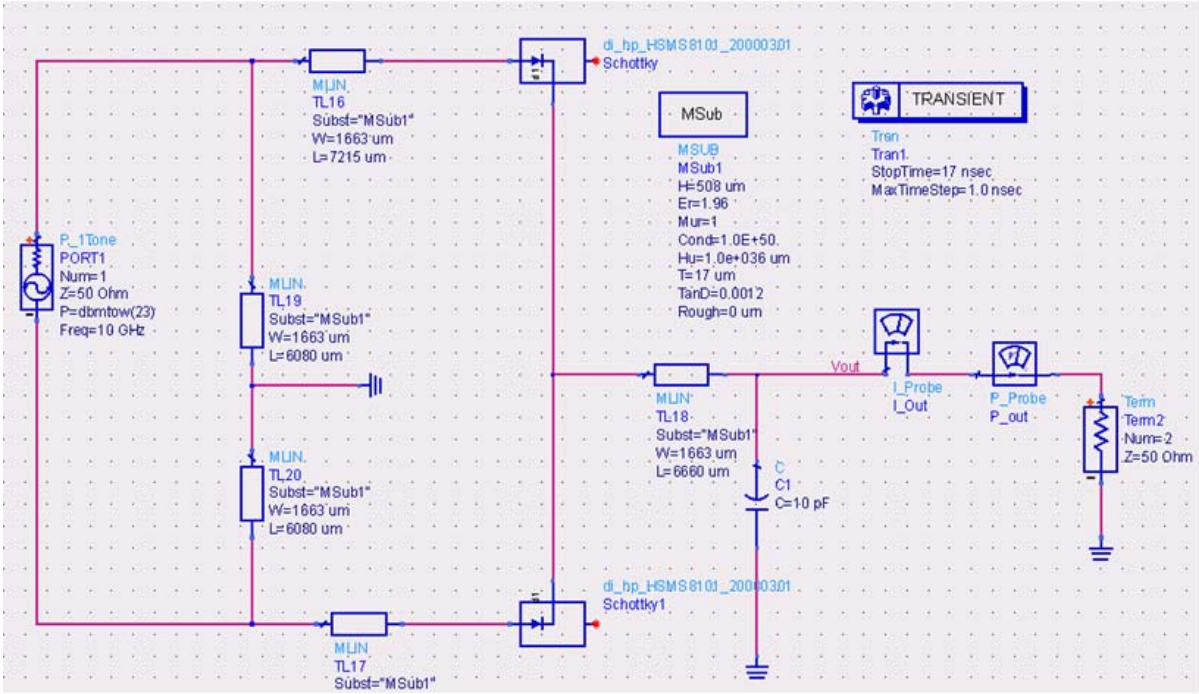

Figure 57. Final circuit design of full-wave rectification rectenna.

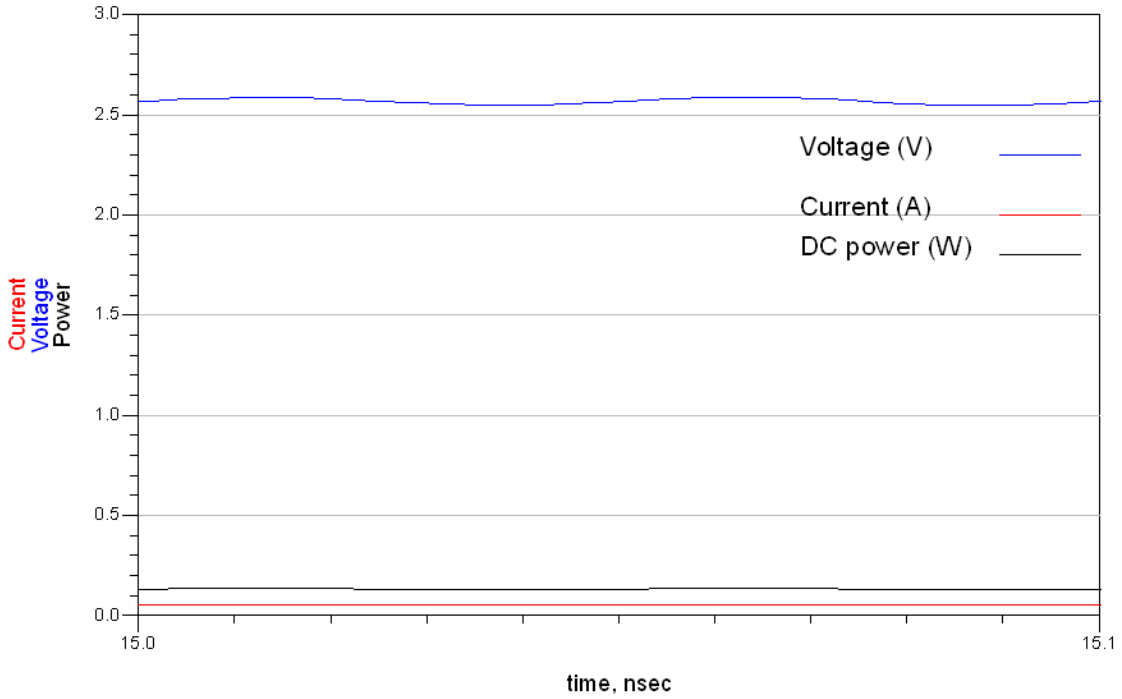

Figure 58. Simulated output power, voltage, and current-versus-time for the full-wave rectenna with a 23 dBm input.

2. Performance of Full-Wave Rectenna Impedance Design

The simulated conversion efficiency of the full-wave rectenna as a function of input power is shown in Figure 59. The full-wave design is able to convert microwave power to dc power with an efficiency of 65.9% at an input power of 200 mW. The simulated output power of the full-wave rectenna design is shown in Figure 60. The full-wave design is able to produce 132 mW at an input power of 200 mW. In addition, the full-wave input–output characteristic is nearly linear, from 20 mW to 200 mW. From 200 mW to 400 mW of input power, the output begins to saturate. As a result, we prefer using the full-wave rectenna design for MAV application at an input power of 200 mW because it achieves the highest efficiency. However, to obtain this high power output requires a very large power density at the antenna, as observed in Figure 19.

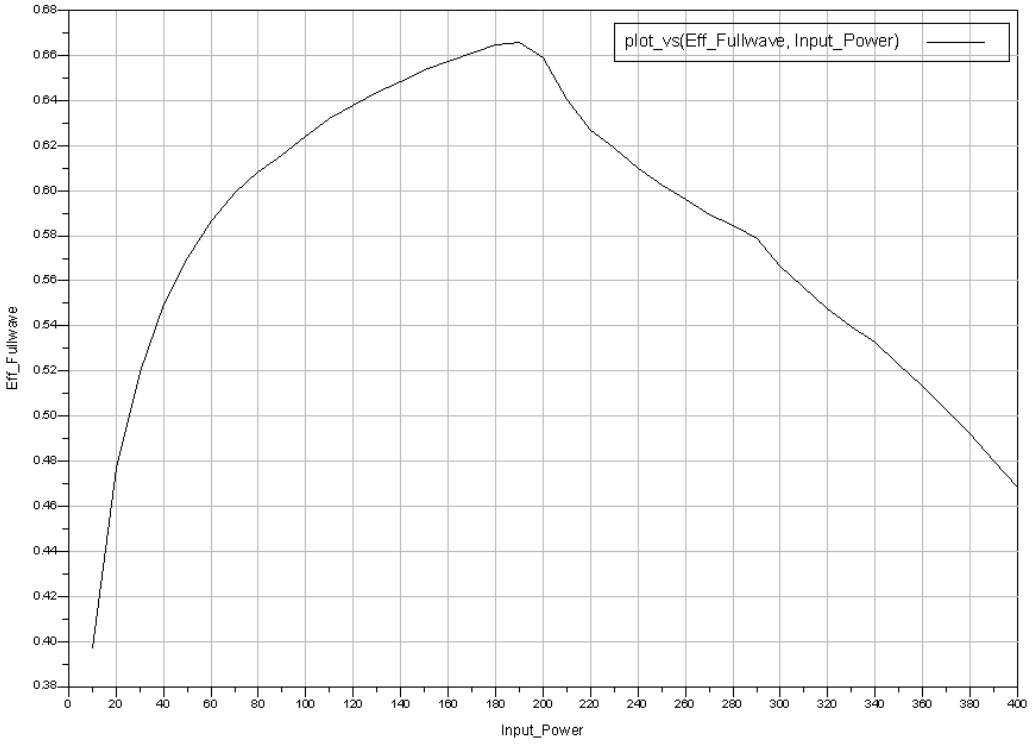

Figure 59. Conversion efficiency of final full-wave rectenna design.

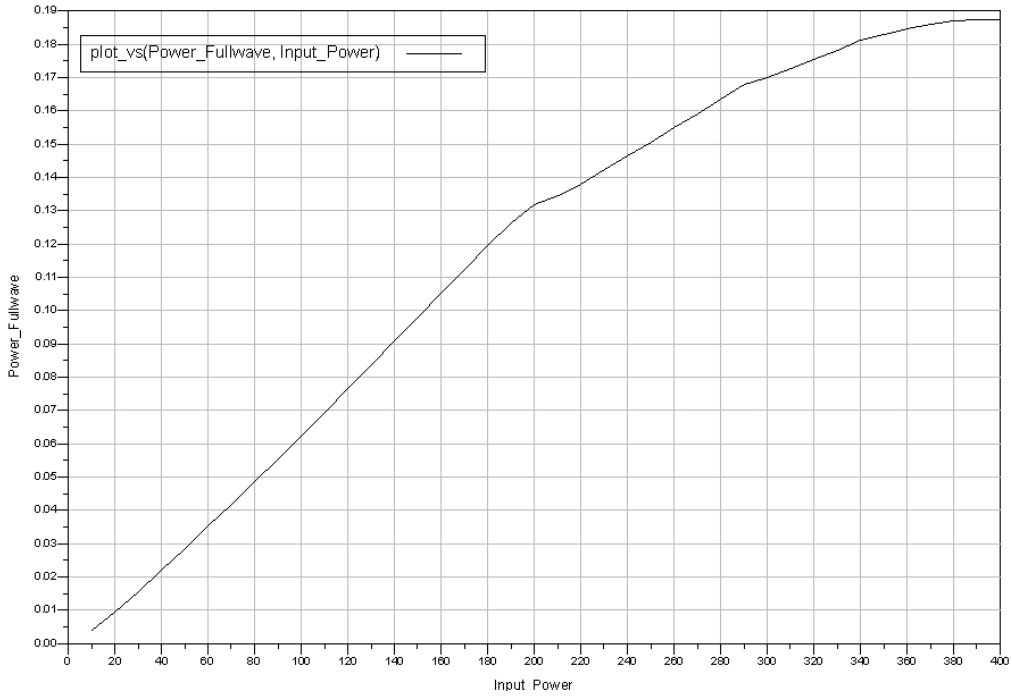

Figure 60. Output power (watts) of final full-wave rectenna design.

B. CIRCUIT LAYOUT AND TUNING

1. Circuit Design Layout

The final circuit-design layout is shown in Figure 61. In Figure 61, the full-wave rectenna includes a matching unit, two diodes, and one capacitor. Based on the ADS simulation, the HSMS 8101 has an S_{22} characteristic impedance of 4.7 + j40 Ω. In order to have a compact circuit design to reduce the size and weight of the printed circuit board (PCB), we bend the impedance-matching unit. The length of ℓ and d are the same values as those in the ADS model in Figure 57. The calculation of the length of L_0 is discussed in the next section. For lightweight design, we select a dielectric constant of 1.96 for the substrate. The properties of the dielectric material are listed in Table 9.

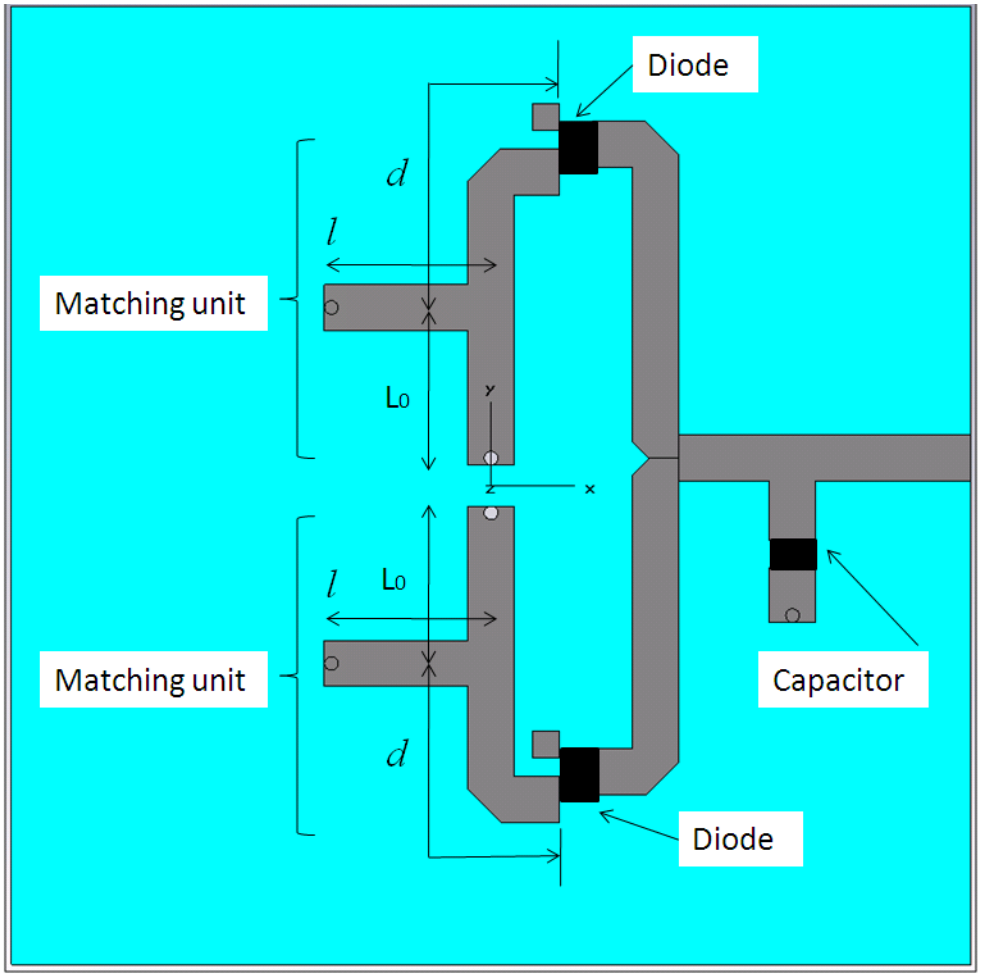

Figure 61. Final circuit layout of the full-wave rectenna modeled in CST.

Dielectric Material	Duroid 5880LZ from Rogers Corporation
Dielectric Constant ε_r	1.96 ± 0.04
Loss Tangent, $\tan\delta$	0.0019 @ 10 GHz
Substrate Height, h	0.508 mm microstrip circuit
Copper Thickness	0.5 oz or 17 μm

Table 9. Properties of Rogers Duroid 5880LZ.

2. Calculation of the Matching Unit

Figure 62 shows the portion of the circuit between an antenna terminal and the diode. As seen in Figure 61, the circuit occurs at both dipole terminals. When a sinusoidal voltage is applied at the terminals, one diode will conduct, appearing as a complex impedance. The values of ℓ and d have been chosen to match the conducting diode to the 50 Ω characteristic impedance of the line. This was done using the ADS parameter sweep.

In order for the dipole to be matched at the conducting diode side, the other side must present an open circuit at the antenna terminal. Therefore L_0 must be chosen so that the combination of the open circuit at the diode (d) and its matching stub (ℓ) present an open circuit at the antenna terminal.

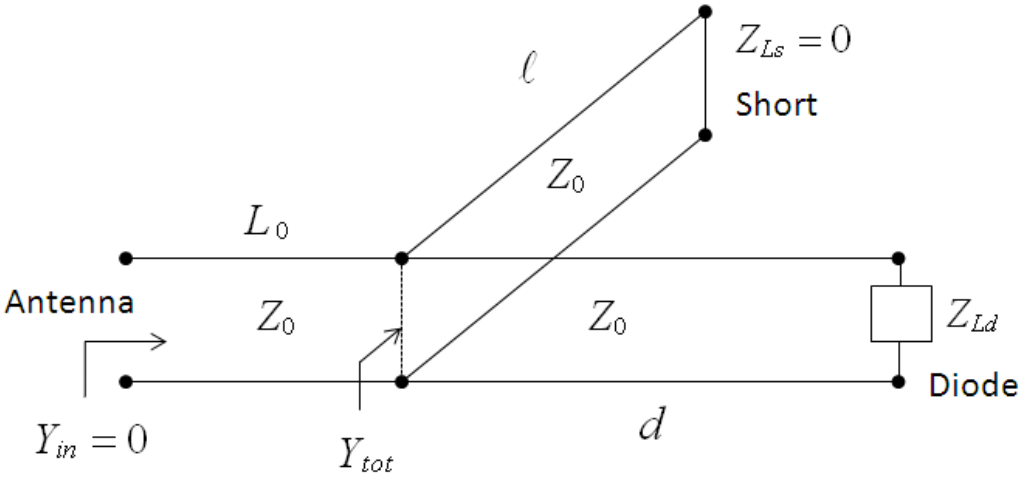

Figure 62. Circuit diagram showing the stub position.

The impedance of the shorted transmission line ℓ at the junction can expressed as

$$Z_{ins} = Z_0 \frac{Z_{Ls} + jZ_0 \tan(\beta\ell)}{Z_0 + jZ_{Ls} \tan(\beta\ell)} = jZ_0 \tan(\beta\ell) \tag{4.1}$$

and the admittance is

$$Y_{ins} = \frac{1}{Z_{ins}} = \frac{1}{jZ_0 \tan(\beta\ell)} \tag{4.2}$$

where Z_0 is 50 Ω and Z_{Ls} is approximately 0 Ω. In addition, the impedance of the transmission line d at the junction can be expressed as

$$Z_{ind} = Z_0 \frac{Z_{Ld} + jZ_0 \tan(\beta d)}{Z_0 + jZ_{Ld} \tan(\beta d)} = \frac{Z_0}{j \tan(\beta d)} \tag{4.3}$$

where

$$Z_{Ld} = \begin{cases} 4.7 + j40 & \text{conducting} \\ \infty & \text{nonconducting} \end{cases} \tag{4.4}$$

and its admittance at the junction is

$$Y_{ind} = \frac{1}{Z_{ind}} = \frac{j \tan(\beta d)}{Z_0}. \tag{4.5}$$

The total admittance at the junction is

$$Y_{tot} = Y_{ins} + Y_{ind} = \frac{-j}{Z_0 \tan(\beta \ell)} + \frac{j \tan(\beta d)}{Z_0}. \tag{4.6}$$

The impedance looking into the junction must be transformed through a distance L_0.

$$Z_{in} = Z_0 \frac{Z_{tot} + jZ_0 \tan(\beta L_0)}{Z_0 + jZ_{tot} \tan(\beta L_0)} = \frac{1}{y_0} \frac{1 + jy_{tot} \tan(\beta L_0)}{y_{tot} + j \tan(\beta L_0)} \tag{4.7}$$

where $y_{tot} = \frac{Y_{tot}}{Y_0}$. Therefore

$$Y_{in} = \frac{1}{Z_{in}} = \frac{y_{tot} + j \tan(\beta L_0)}{1 + jy_{tot} \tan(\beta L_0)}. \tag{4.8}$$

The maximum power from the antenna occurs when Y_{in} equals to 0, therefore

$$y_{tot} + j \tan(\beta L_0) = 0 \tag{4.9}$$

or

$$L_0 = \frac{1}{\beta} \tan^{-1}\left(j \frac{Y_{tot}}{Y_0} \right). \tag{4.10}$$

This is the initial value of the length of the stub L_0 which is a function of Y_{tot}. The length of this stub depends on the length of matching lines ℓ and d.

The parameter sweep in CST was used to fine-tune the value of L_0. Figure 63 shows the CST model of the circuit with one terminal of the dipole connected to a 50 Ω line and the other diode open, to simulate a nonconducting half cycle. Figure 64 shows the best match at 10 GHz obtained with the parameter sweep. The value of L_0 is 4.226 mm, which was used in the final design. The radiation pattern for the circuit of Figure 63 is shown in Figure 65.

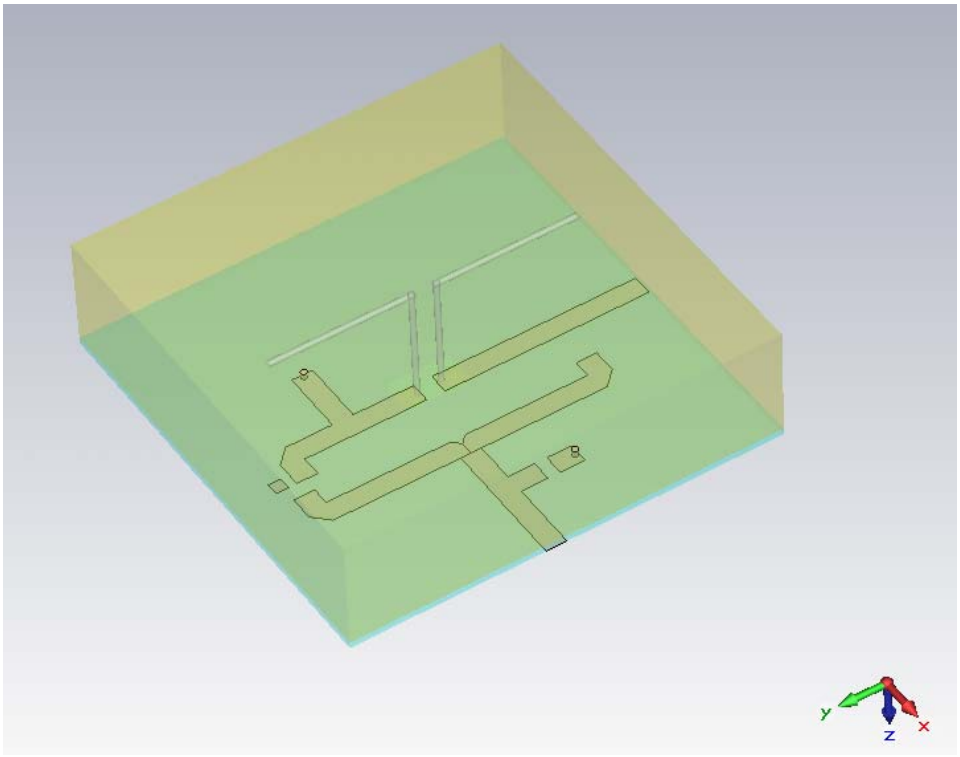

Figure 63. CST model of the circuit.

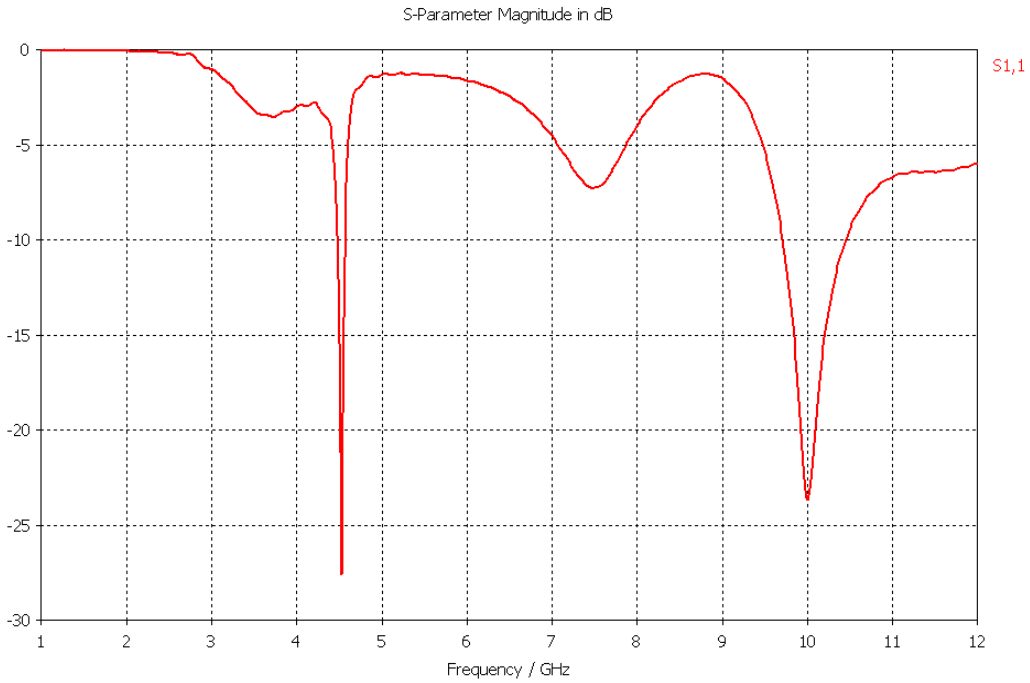

Figure 64. S_{11} frequency response of full-wave rectenna using CST.

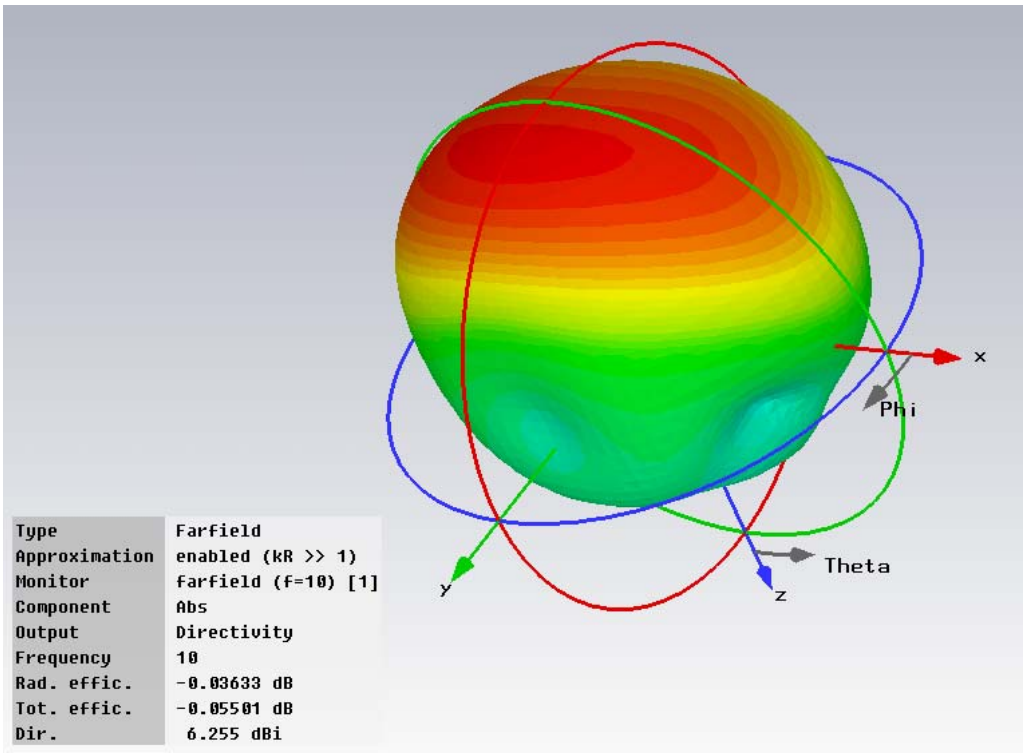

Figure 65. Simulated result of the far-field radiation of full-wave rectenna using CST.

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V. SUMMARY, CONCLUSIONS AND RECOMMENDATIONS

A. SUMMARY

This thesis evaluated different rectenna designs for wireless power transmission for MAV applications using the ADS simulation software. The first and second chapters covered the history of WPT and research performed previously at NPS. Various rectification circuit designs and antennas were simulated using Agilent ADS software and CST Microwave Studio. A shorted stub microstrip tuner was introduced for impedance matching. The dipole-antenna array was selected to reduce the weight of the rectenna in order to hover a MAV.

A full-wave rectenna without low-pass filter was selected to reduce the weight of the rectenna element yet achieve high conversion efficiency. In addition, a harmonic balance analysis was conducted in order to compare the influence of the harmonics on various circuit designs. The full-wave rectenna circuit was simulated to verify performance, and the conversion efficiency is about 65%. The length of matching units were calculated and optimized in Microwave Studio in order to lay out the circuit.

B. RECOMMENDATIONS

The result of this thesis shows that the improved full-wave rectanna converts more dc power than a half-wave rectenna, as verified by simulation. A dipole antenna was used instead of a circular-patch antenna in order to reduce the weight of the rectenna system. The conversion efficiency of the rectenna can be improved, however, and some recommendations follow.

1. Using High-Power Transmitter and High-Gain Antenna

The performance of the WPT can be further enhanced by complementing it with a high-power transmitter and high-gain antenna to increase the power-density incident on the rectenna array.

2. Building a Hardware Prototype

Upon the successful design of the dipole antenna and rectifier circuit using CST Microwave Studio, it is necessary to build a prototype of the full-wave rectenna to measure the output dc voltage and current and determine the realized efficiency.

3. Running Simulations for an Array

The MWS simulation should be done for an array of dipoles to assess the effect of mutual coupling. The array element may need to be returned to compensate for the mutual coupling.

4. Reducing Polarization Loss

The dipole is linearly polarized, and therefore loss of signal will occur if the MAV antenna is orientated in a cross-polarized direction in flight. A circularly polarized antenna should be considered to reduce the polarization loss factor.

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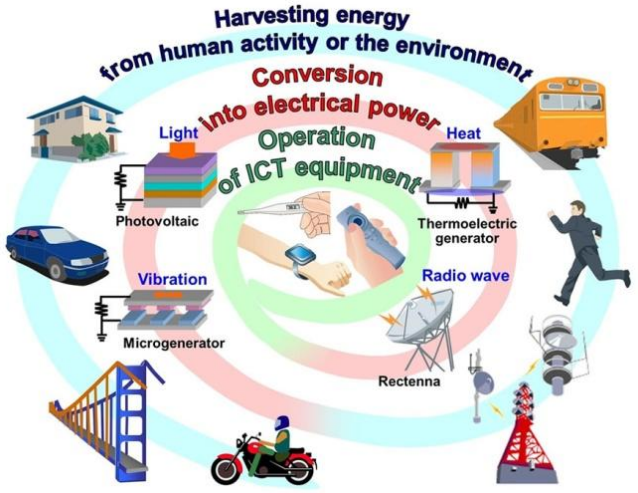
Energy Harvesting

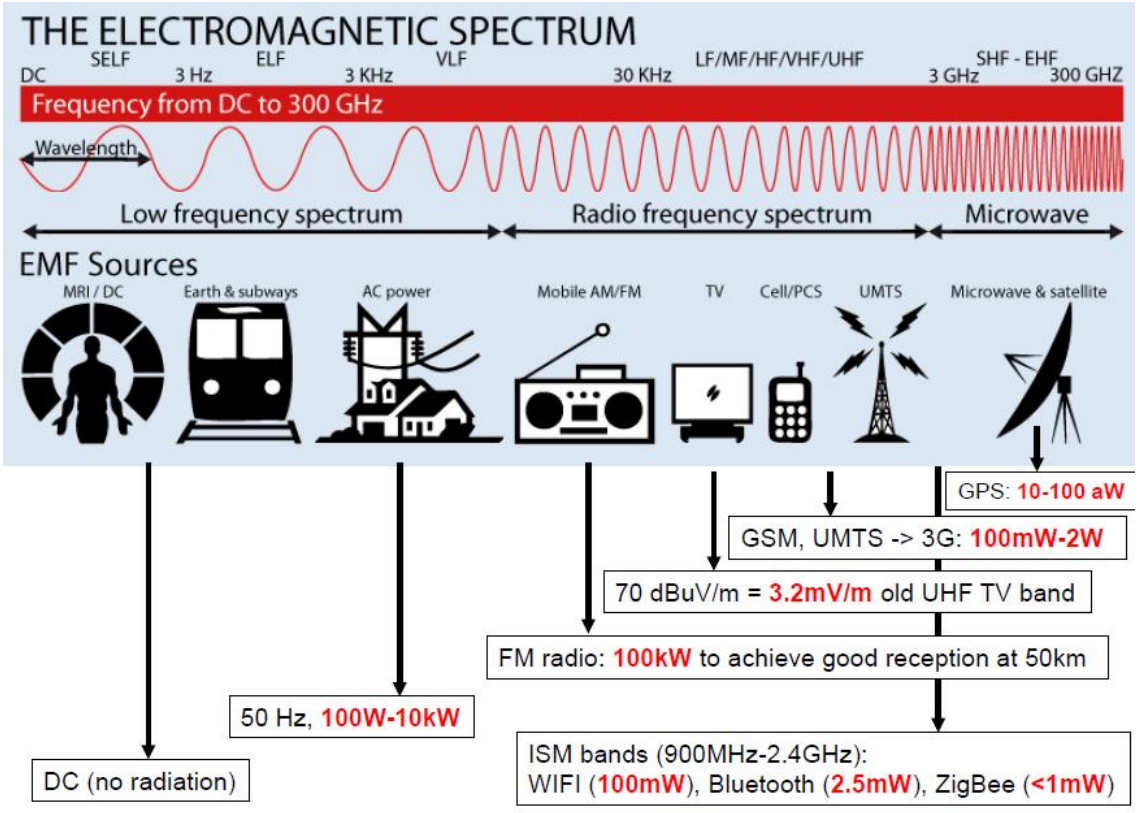
Energy Harvesting

Energy harvesting is the process by which energy is derived from external sources (e.g. solar power, thermal energy, wind energy, salinity gradients, and kinetic energy), captured, and stored for small, wireless autonomous devices.

Energy harvesters provide a very small amount of power for low-energy electronics. The energy source for energy harvesters is present as ambient background and is free. For example, temperature gradients exist from the operation of a combustion engine and in urban areas, there is a large amount of electromagnetic energy in the environment because of radio and television broadcasting.



Energy Harvesting



Energy Harvesting

- Examples of small-scale energy harvesters are:
- Wristwatches powered by kinetic energy (called automatic watches) where the movement of the arm is used. The arm movement causes winding of its mainspring. A newer design uses movement of a magnet in the electromagnetic generator instead to power the quartz movement. The motion provides a rate of change of flux, which results in some induced EMF on the coils. The concept is simply related to Faraday's Law.
- Photovoltaics is a method of generating electrical power by converting solar radiation (both indoors and outdoors) into direct current electricity using semiconductors that exhibit the photovoltaic effect. Conventional photovoltaic power generation employs solar panels composed of a number of cells containing a photovoltaic material.
- Thermoelectric generators consist of the junction of two dissimilar materials and the presence of a thermal gradient. Large voltage outputs are possible by connecting many junctions electrically in series and thermally in parallel. Typical performance is 100-200 μV/K per junction. These can be utilized to capture mW of energy from industrial equipment, structures, and even the human body. They are typically coupled with heat sinks to improve temperature gradient.

Energy Harvesting

- Micro wind turbine are used to harvest wind energy readily available in the environment in the form of kinetic energy to power the low power electronic devices such as wireless sensor nodes. When air flows across the blades of the turbine, a net pressure difference is developed between the wind speeds above and below the blades. This will result in a lift force generated which in turn rotate the blades.
- Piezoelectric crystals or fibers generate a small voltage whenever they are mechanically deformed. Vibration from engines can stimulate piezoelectric materials, as can the heel of a shoe, or the pushing of a button.
- Special antennas can collect energy from stray radio waves, this can also be done with a Rectenna and theoretically at even higher frequency EM radiation with a Nantenna.
- Even power from keys pressed during use of a portable electronic device or remote controller, using magnet and coil or piezoelectric energy converters, may be used to help power the device.

Energy Harvesting

- Low power has been the most important electronic design criterion for at least the last ten years. Thanks to Moore's Law, semiconductor power levels have dropped dramatically, often consuming milliwatts in run mode and nanowatts in standby mode. As a direct result, ultra-low-power wireless sensorless networks finally became possible and their adoption has been dramatic. Now, sensors stand alone in remote or hard-to-reach areas to warn of building and bridge stresses, air pollution, forest fires, pending landslides, worn bearings, and wing vibration. Low-power wireless sensor networks are at the heart of numerous industrial, medical, and commercial applications.

- However, off-grid, as well as portable sensor nodes, rely on batteries for power and face the same problem as cell phones. In such cases, it is advisable to prolong battery life by harvesting environmental energy sources – most often available as light, heat, vibration, motion, or ambient RF. If a device's energy requirement is low enough and battery replacement would be difficult or expensive, it may be possible to rely exclusively on harvesting ambient energy sources for power. The combination of ultra-low-power MCUs and energy harvesting have given rise to a wealth of applications that previously were not possible.

Energy Harvesting

- The energy harvesting market is large and growing rapidly. According to analysts, energy harvesting was a $0.7 billion market in 2012 and is expected to exceed $5 billion by 2022; by then 250 million sensors will be powered by energy harvesting sources. The market for thermoelectric energy harvesting alone will reach $865 million by 2023.

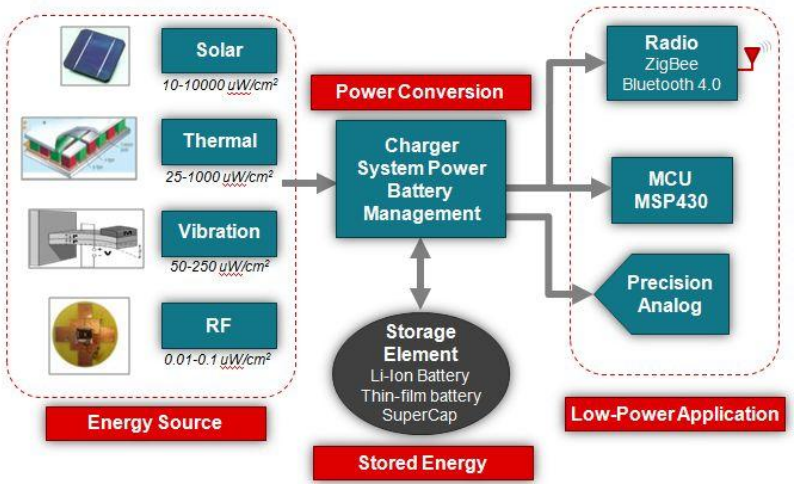
Source	Source Power	Harvested Power
Light		
Indoor	0.1 mW/cm²	10 µW/cm²
Outdoor	100 mW/cm²	10 mW/cm²
Vibration/Motion		
Human	0.5m at 1 Hz	
	1m/s² at 50 Hz	4 µW/cm²
Machine	1m at 5 Hz	
	10m/s² at 1 kHz	100 µW/cm²
Thermal		
Human	20 mW/cm²	30 µW/cm²
Machine	100 mW/cm²	1-10 mW/cm²
RF		
GSM BSS	0.3 µW/cm²	0.1 µW/cm²

Energy Harvesting

Solar

There is hardly a home or office that does not have at least one solar-powered calculator – actually, a calculator with a coin-cell battery and a small front panel photovoltaic (PV) cell to top it up. These polycrystalline silicon or thin-film cells convert photons to electrons with a typical efficiency of about 15 to 20% for polycrystalline and 6 to 12% for thin film cells. Since the power available from indoor lighting is typically only about 10 µW/cm², their usefulness depends on the size of the module plus the spectral composition of the light.

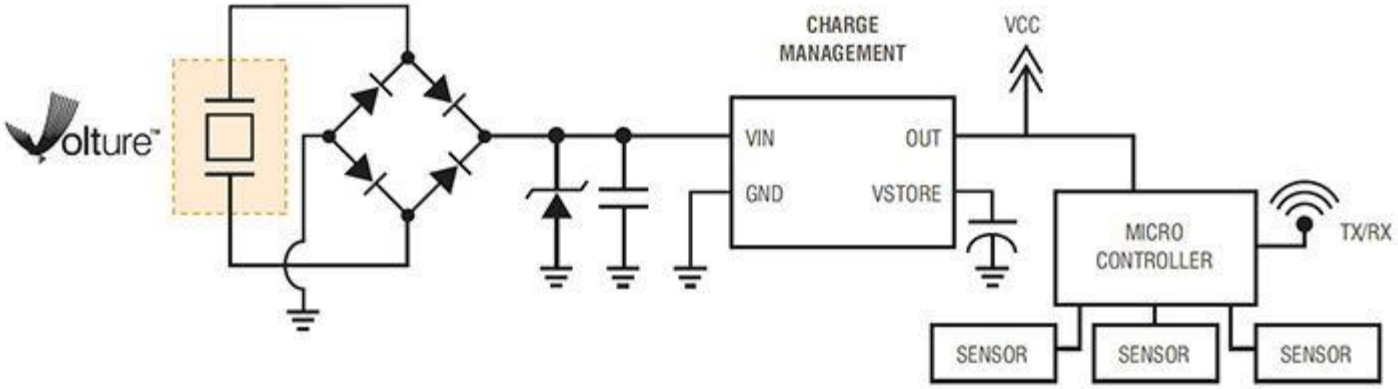
Small solar cells are frequently used in consumer and industrial applications, including toys, watches, calculators, street lighting controls, portable power supplies, and satellites. Since light sources tend to be intermittent, solar cells are used to charge batteries and/or supercapacitors to provide a stable energy source.



Energy Harvesting

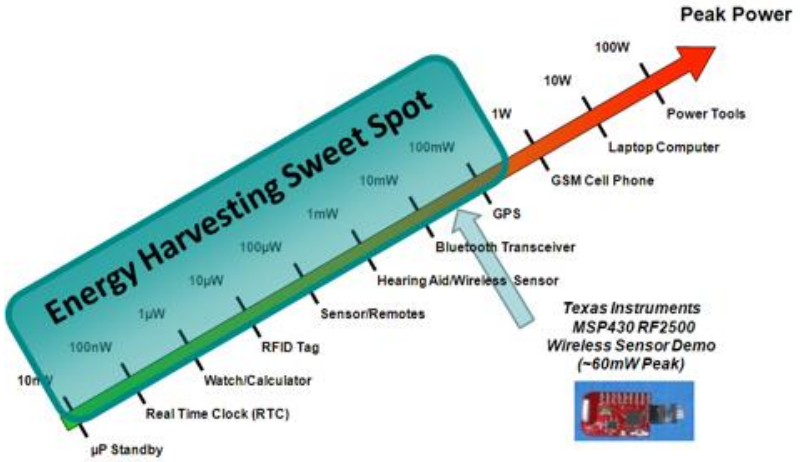
Piezoelectric

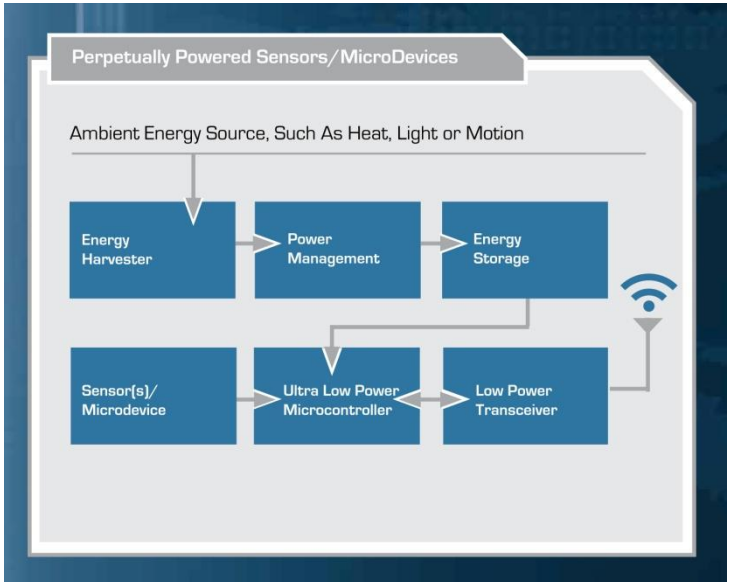
Piezoelectric transducers generate electricity when stressed, which make them good candidates for vibration sensors when they are used in energy harvesting modules that detect motor bearing noise and the vibration of aircraft wings. The Midé Volture™ V-20W Vibration Energy Harvester employs a cantilever that attaches to a piezoelectric crystal. When vibrations set the cantilever in motion it generates an AC output voltage that is rectified, regulated, and stored in a supercapacitor or thin-film battery.



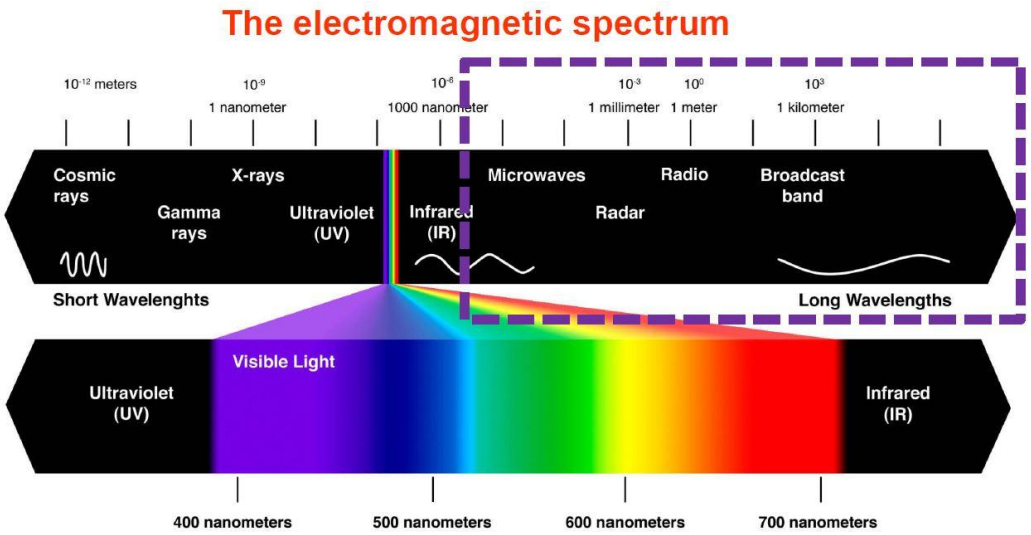
Midé Volture™ piezoelectric energy harvester (Courtesy of Midé)

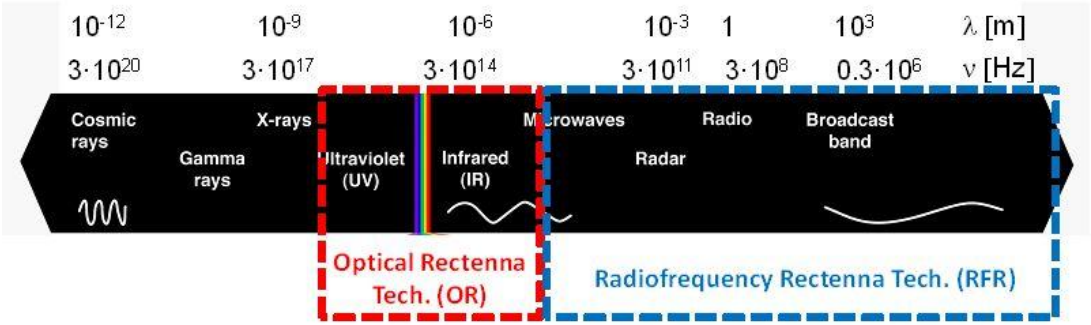
Energy Harvesting





RF Energy Harvesting





RF Energy Harvesting

Radio Frequency - RF

RFID works by rectifying a strong local signal aimed directly at the sensor.

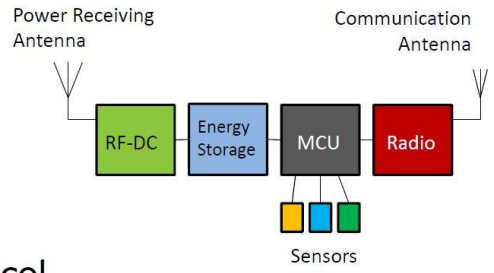
When harvesting RF ambient energy it must be taken into account that RF energy is generally very low, therefore:

– Direct-power at close range to a transmitter

– Energy accumulation for longer range

are needed. Simple battery-recharge is possible but a system level approach is required for optimal implementation. Key system elements are reported below:

- Harvester
- Antenna
- Storage
- MCU
- Sensors
- Radio
- Comm. Protocol



RF Energy Harvesting

System Parameters

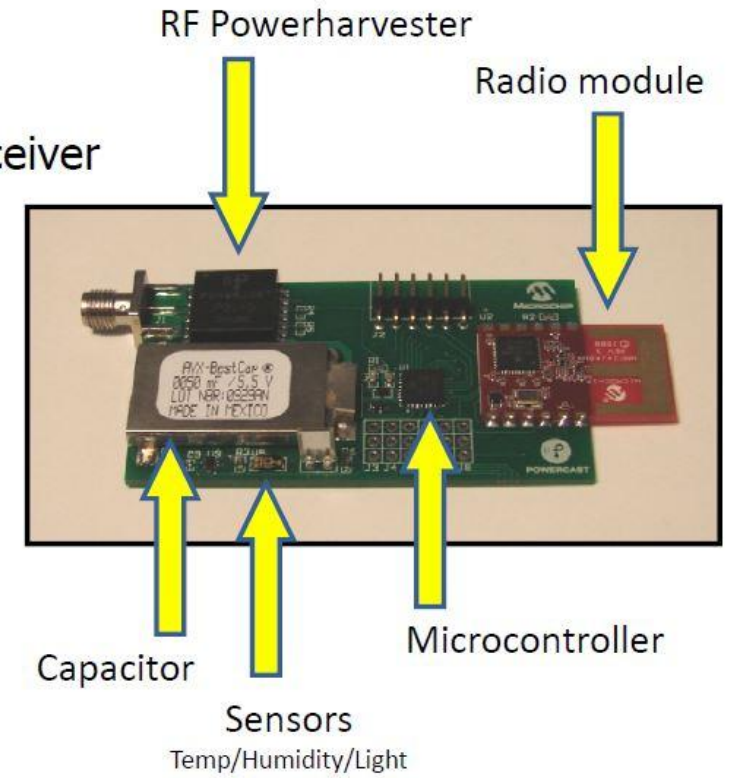
- Operation
  - continuous vs. intermittent
- RF power source
  - distance, power, frequency
- Receiving antenna
  - size, performance (gain)
- Operating voltage
- Energy Storage
- Sensors (active, passive)

RF Energy Harvesting

Designed for Low-Power RF Harvesting

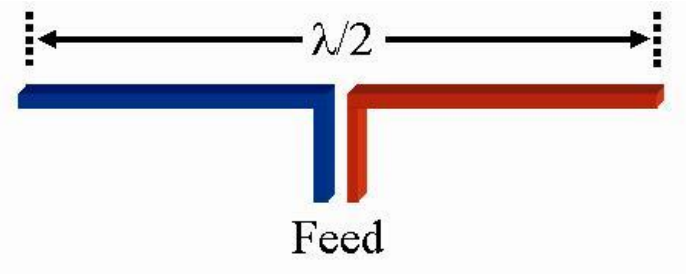
Module Components & Features

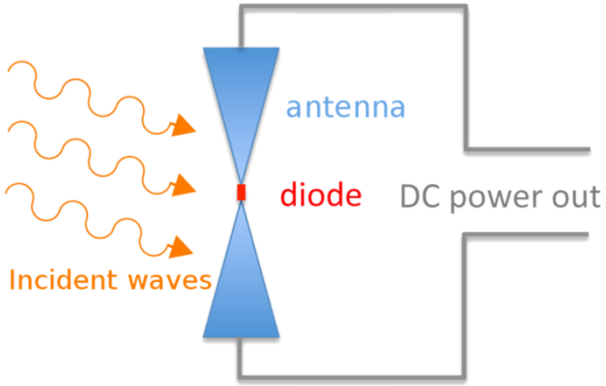
- Powercast P2110 Powerharvester™ Receiver
- MCU: Microchip PIC24 XLP
- Radio module: Microchip MRF24 (802.15.4)
- System power: 3.3V
- Capacitor: 50mF (as low as 3300uF)
- Discrete sensors: Temp, Humidity, Light
- Wireless protocol: MiWi™ P2P



RF Energy Harvesting

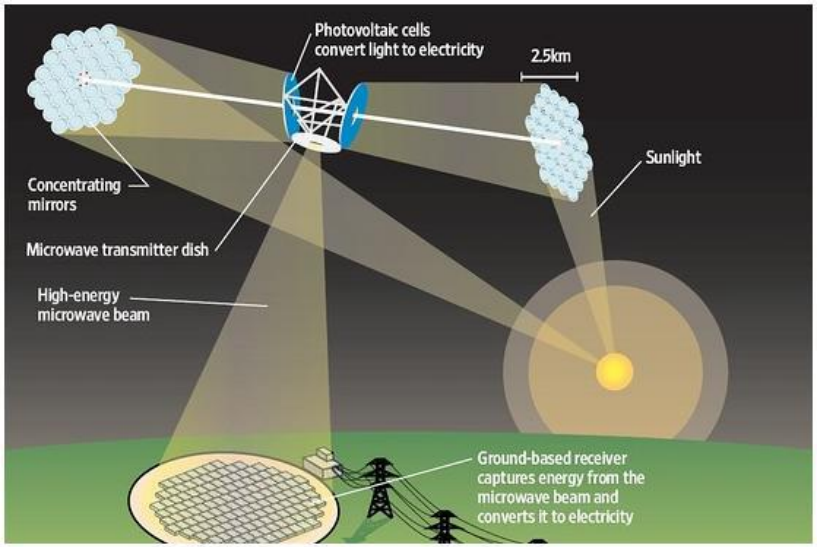
- The RF energy harvester is typically called **rectenna**, i.e. **rect**ifying ant**enna**, a special type of antenna that is used to convert electromagnetic energy into direct current electricity. This component is used in wireless power transmission systems that harvest power by radio waves. A simple rectenna element consists of a dipole antenna with an RF diode connected across the dipole elements. The diode rectifies the AC current induced in the antenna by the microwaves, to produce DC power, which powers a load connected across the diode. Schottky diodes are usually used because they have the lowest voltage drop and highest speed and therefore have the lowest power losses due to conduction and switching. Large rectennas consist of an array of many such dipole elements.





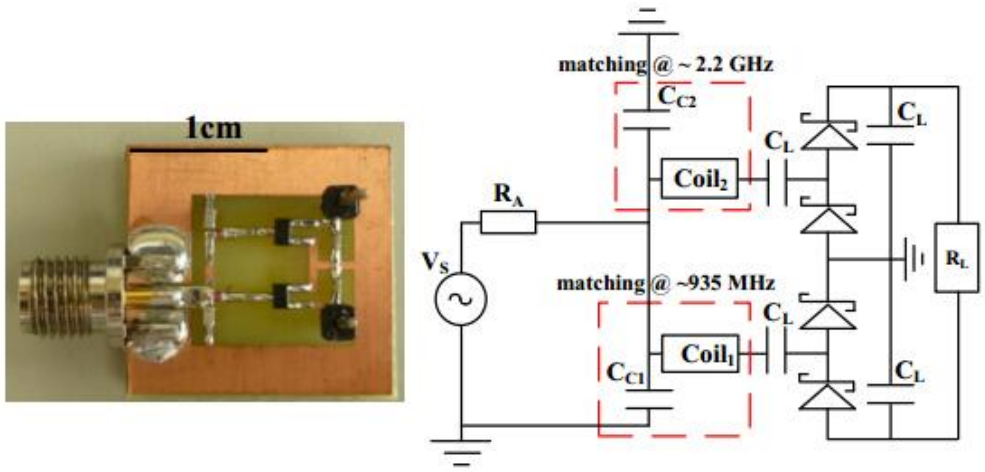
RF Energy Harvesting

- The invention of the rectenna in the 1960s made long distance wireless power transmission feasible. The rectenna was invented in 1964 and patented in 1969 by US electrical engineer William C. Brown, who demonstrated it with a model helicopter powered by microwaves transmitted from the ground, received by an attached rectenna. Since the 1970s, one of the major motivations for rectenna research has been to develop a receiving antenna for proposed solar power satellites, which would harvest energy from sunlight in space with solar cells and beam it down to Earth as microwaves to huge rectenna arrays. A proposed military application is to power drone reconnaissance aircraft with microwaves beamed from the ground, allowing them to stay aloft for long periods.



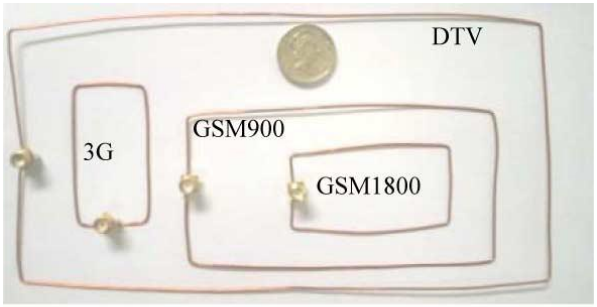
RF Energy Harvesting

Now the interest has turned to using rectennas as power sources for small wireless microelectronic devices. The largest current use of rectennas is in RFID tags, proximity cards and contactless smart cards, which contain an integrated circuit (IC) which is powered by a small rectenna element. When the device is brought near an electronic reader unit, radio waves from the reader are received by the rectenna, powering up the IC, which transmits its data back to the reader.



Picture and circuit layout of a dual-band RF harvester.
The harvester is matched at 935MHz and 2.2GHz.

RF Energy Harvesting



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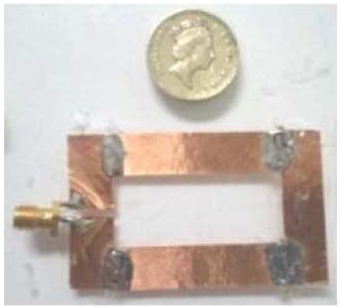

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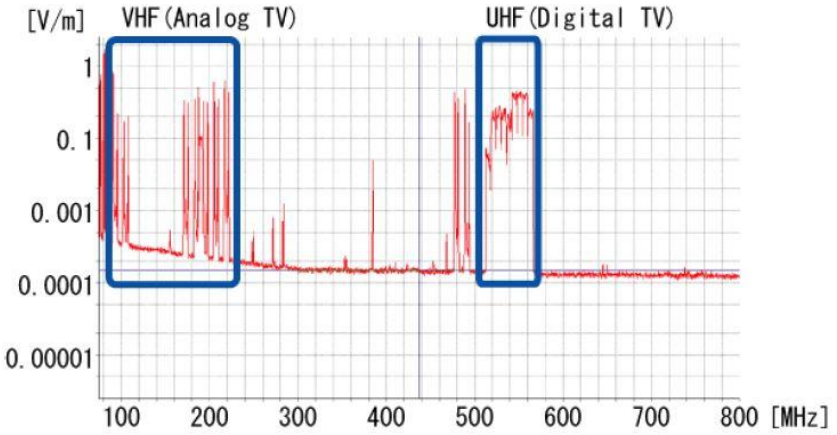
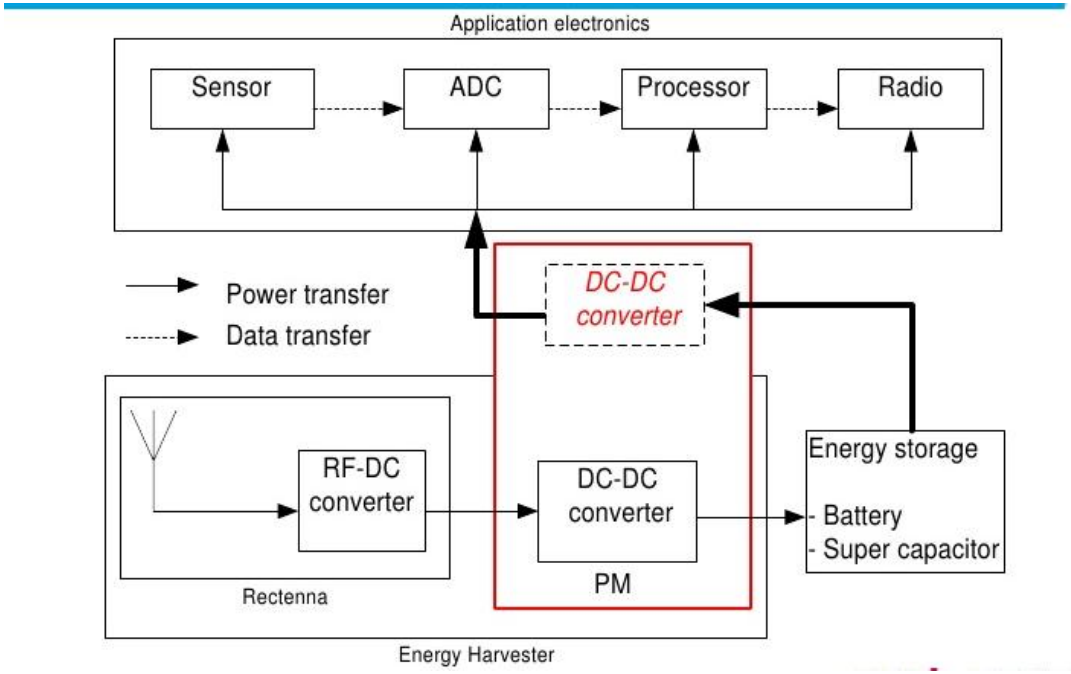


Field intensity in the area surrounding the Tokio tower (6.6 km distance apart) due to TV signals.

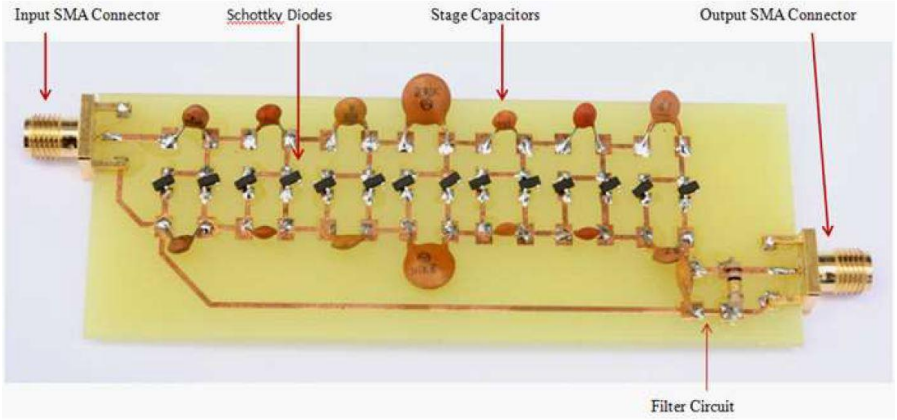
Folded-dipole antennas shown next to a British £1 coin. (a) DTV, GSM900, GSM 1800 and 3G copper wire antennas. (b) 3G copper tape antenna on Perspex.

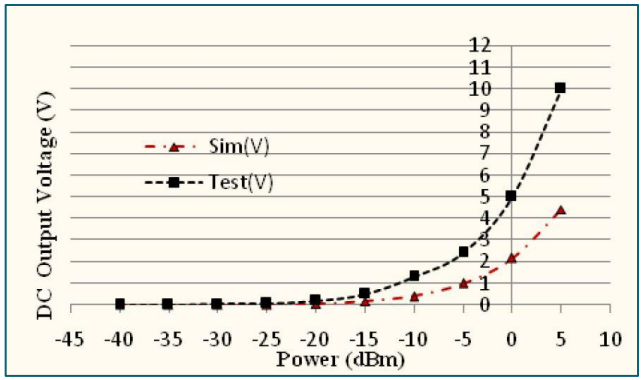
RF Energy Harvesting



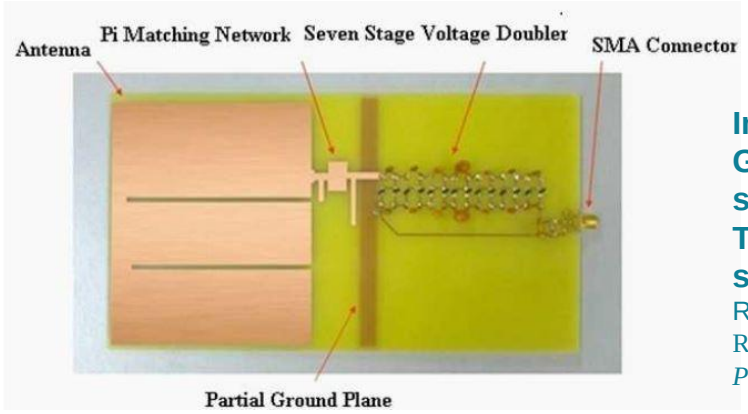
Block diagram of an RF harvesting system

RF Energy Harvesting





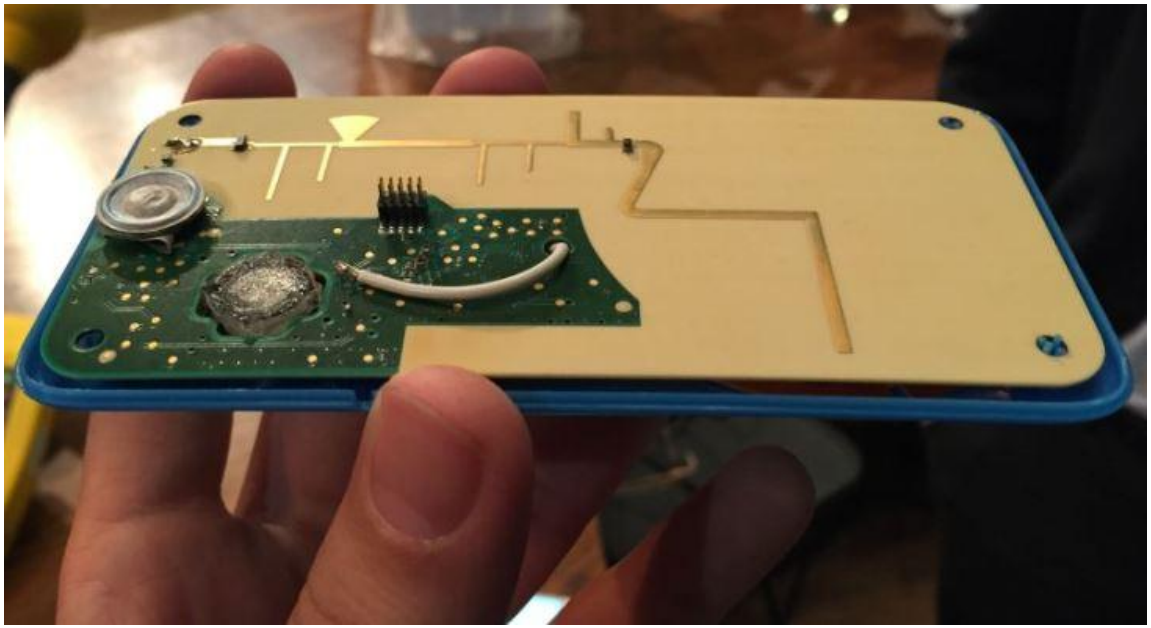
Assembled circuit board of a 7-stage voltage doubler in the RF-DC conversion module and performance test



Integrated assembled PCB of an RF energy harvesting system for GM-900 signals. The DC voltage obtained from the harvester system in the field test at 50m from a GSM cell tower was 2.9 V. This voltage was enough to power an STLM20 temperature sensor.

Ref. N. M. Din, C. K. Chakrabarty, A. Bin Ismail, K. K. A. Devi, W.-Y. Che, "DESIGN OF RF ENERGY HARVESTING SYSTEM FOR ENERGIZING LOW POWER DEVICES", *Progress In Electromagnetics Research, Vol. 132, 49{69, 2012*]

RF Energy Harvesting



The bulk of this Freevolt unit is the antenna designed for ambient RF energy harvesting from high-power broadcast (DTV) and mobile network (Wi-Fi, GSM, LTE) signals. On the left is a speaker ticking every time enough RF energy is harvested to make it move.

http://arstechnica.co.uk/gadgets/2015/09/freevolt-perpetual-free-rf-energy-harvesting-to-power-the-internet-of-things/

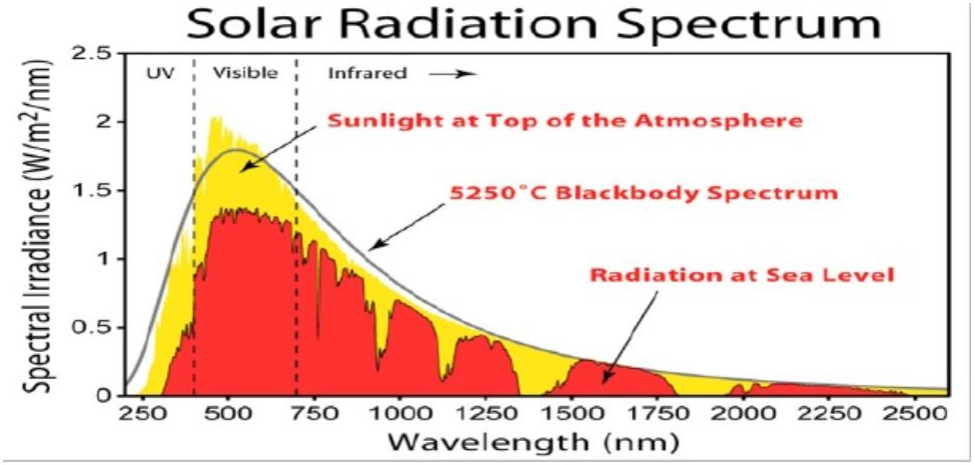
Advanced Solar Harvesting

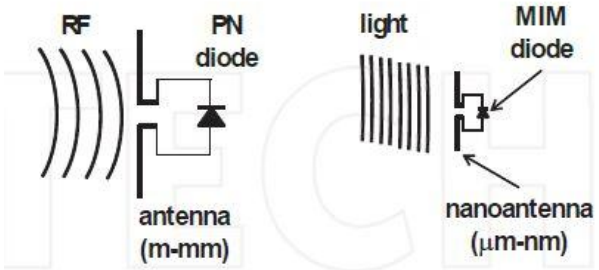
New strategies for solar energy harvesting employing other devices than solar cells, are currently under study. To this end, the nantenna or nano-antenna is the key element for capturing the optical energy. Nano-antennas are resonant metallic structures that confine the optical energy into small volumes in an efficient way by inducing a high-frequency current in its structure. In other words, the nantenna is a very small rectenna (the size of a light wave), fabricated using nanotechnology, which acts as an "antenna" for light, converting light into electricity.

A single nanoantenna reaches a high light-absorbing efficiency if compared to conventional photovoltaic devices. It is hoped that arrays of nantennas could be an efficient means of converting sunlight into electric power, producing solar power more efficiently than traditional solar cell-based systems.

A nantenna is an electromagnetic collector designed to absorb specific wavelengths that are proportional to the size of the nantenna. Currently, **Idaho National Laboratories** has designed a nantenna to absorb wavelengths in the range of 3–15 μm. These wavelengths correspond to photon energies of 0.08-0.4 eV. Based on antenna theory, a nantenna can absorb any wavelength of light efficiently provided that its size is optimized for that specific wavelength. Ideally, nantennas would be used to absorb light at wavelengths between 0.4–1.6 μm because these wavelengths have higher energy than far -infrared (longer wavelengths) and make up about 85% of the solar radiation spectrum.

Advanced Solar Harvesting





Advanced Solar Harvesting

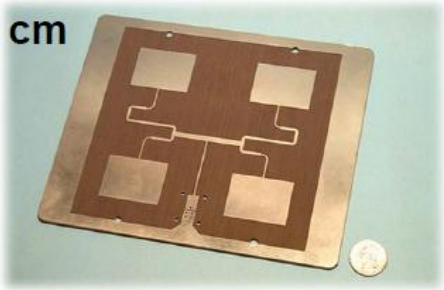
The nanoantennas target mid-infrared rays, which the Earth **continuously** radiates as heat after absorbing energy from the sun during the day. In contrast, traditional solar cells can only use visible light, rendering them idle after dark. Infrared radiation is an especially rich energy source because it also is generated by industrial processes.

Prototype IR nanoantennas have been realized as tiny gold squares set in a specially treated form of polyethylene, thus appearing as light and flexible sheets. Theoretical efficiency as high as 80% has been found and real-life tests are under development. Traditional solar cells rely on a response mechanisnm that only works for up to 20% of the visible light they collect. Scientists have developed more complex solar cells with higher efficiency, but these models are too expensive for widespread use. If technical hurdles on super-fast rectifiers can be overcome, nanoantennas have the potential to be a cheaper, more efficient alternative to solar cells.

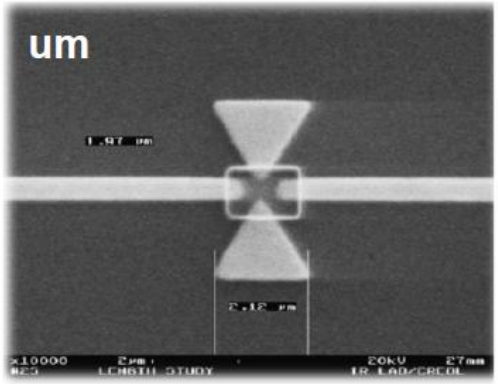
Besides, the nanoantennas also have the potential to act as cooling devices that draw waste heat from buildings or electronics without using electricity. Since objects give off heat as infrared rays, the nanoantennas could collect those rays and re-emit the energy at harmless wavelengths. Such a system could cool down buildings and computers without the external power source required by air-conditioners and fans.

Advanced Solar Harvesting

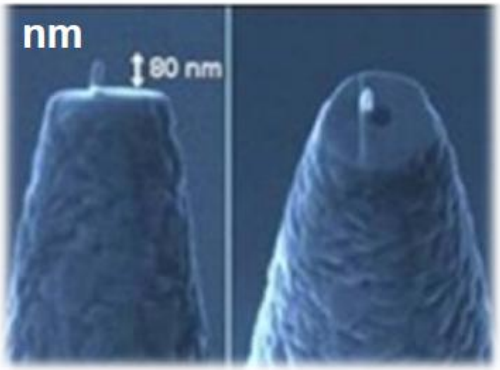
Optical nano-antenna technology



Microwave patch-antenna

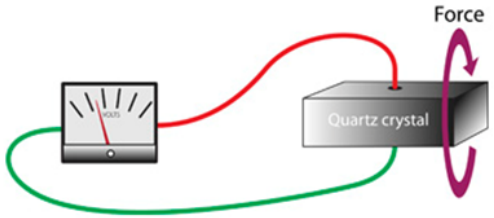


Infrared bow-tie antenna

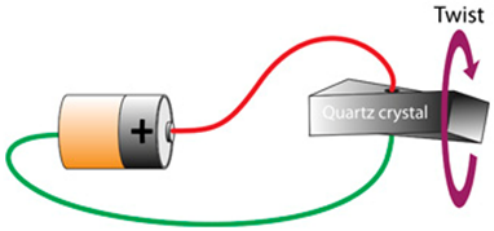


Visible dipole antenna

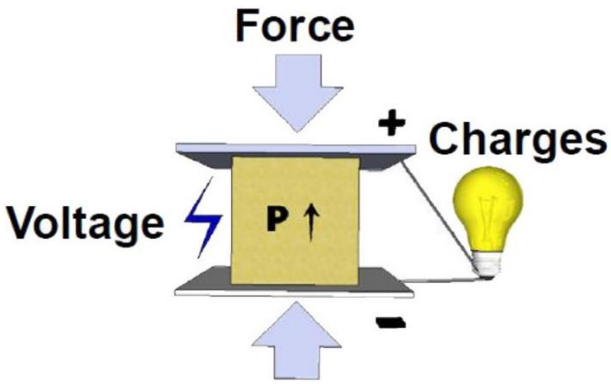
Advanced Piezoelectric Harvesting



Piezoelectricity is the ability of certain materials to produce a voltage when subjected to mechanical stress.



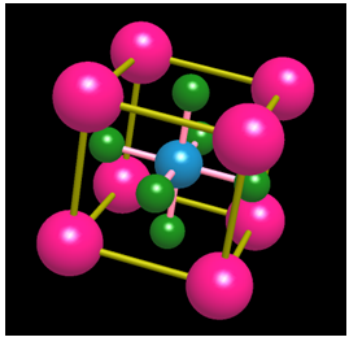
Piezoelectric materials also show the opposite effect, where application of an electrical field creates mechanical stress (size modification) in the crystal.

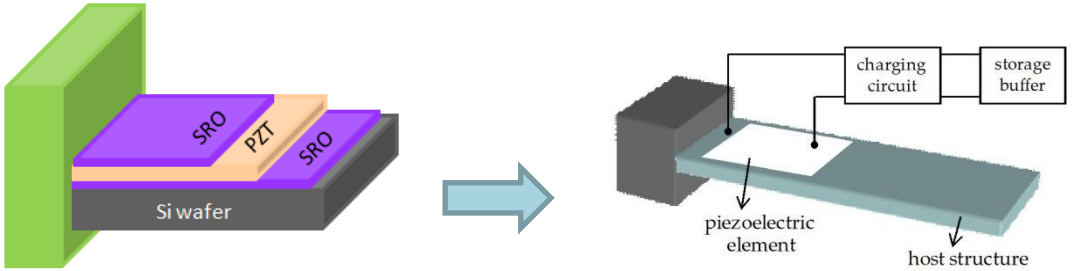


Advanced Piezoelectric Harvesting

Piezo MEMS

- Epitaxial PZT thin films with enhanced piezoelectric properties grow on standard Silicon or SOI wafers
- Compatible with MEMS process, enable production in industry scales.
- Reduced piezo component size lead to revolutionary high density integration

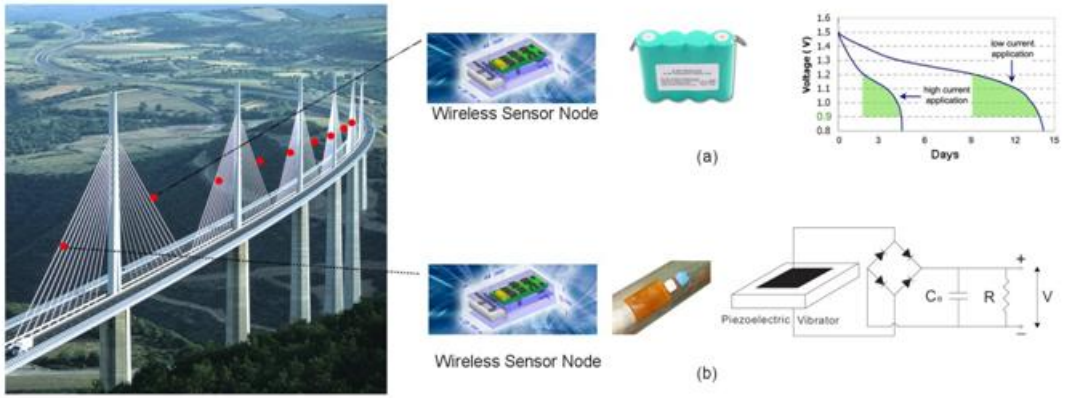




Advanced Piezoelectric Harvesting

Societal Impact

Piezoelectric materials, such as the lead zirconate titanate (PZT), are great candidates for energy harvesting using vibrations from the surrounding environment.

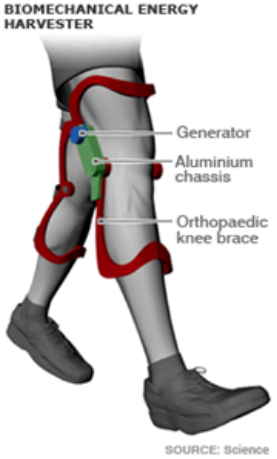


Advanced Piezoelectric Harvesting

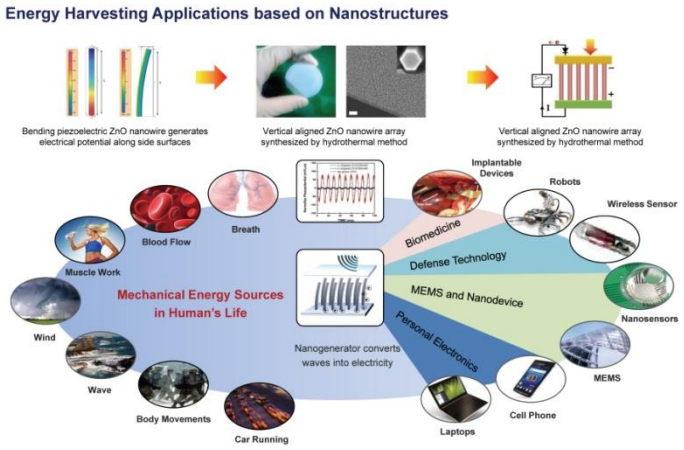
Application Areas

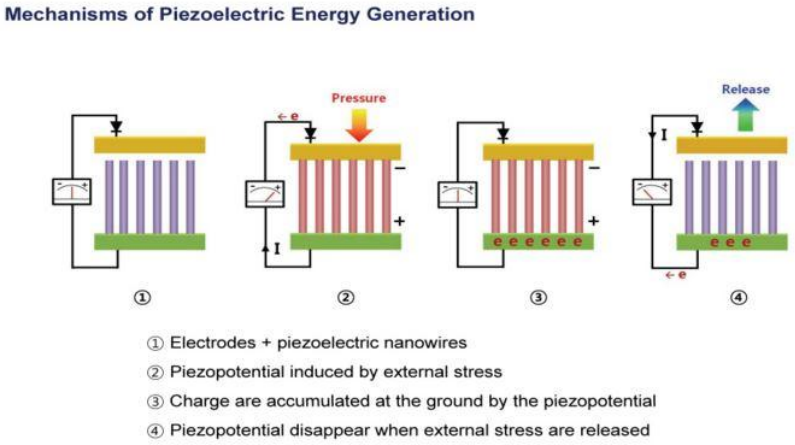
Wireless corrosion monitoring systems

Biomechanical Energy Harvester



Advanced Piezoelectric Harvesting

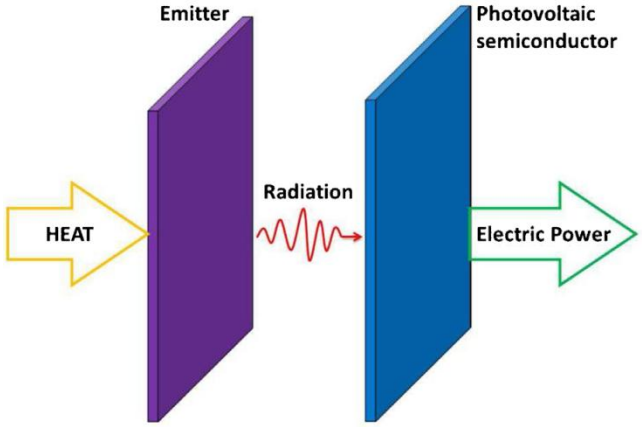




Thermophotovoltaic Harvesting

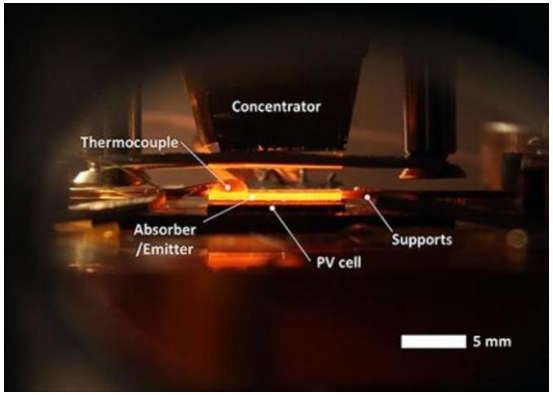
A new approach to harvesting solar energy, developed by MIT researchers, could improve efficiency by using sunlight to heat a high-temperature material whose infrared radiation would then be collected by a conventional photovoltaic cell. A conventional silicon-based solar cell “doesn’t take advantage of all the photons”. This occurs because converting the energy of a photon into electricity requires that the photon’s energy level match that characteristic of the photovoltaic (PV) material called bandgap. Silicon’s bandgap responds to many wavelengths of light, but misses many others.

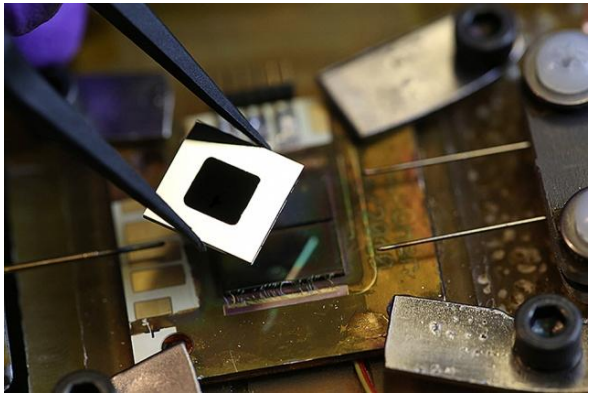
To address that limitation, the research team inserted a two-layer absorber-emitter device — made of novel materials including carbon nanotubes and photonic crystals — between the sunlight and the PV cell. This intermediate material collects energy from a broad spectrum of sunlight, heating up in the process. When it heats up, as with a piece of iron that glows red hot, it emits light of a particular wavelength, which in this case is tuned to match the bandgap of the PV cell mounted nearby.



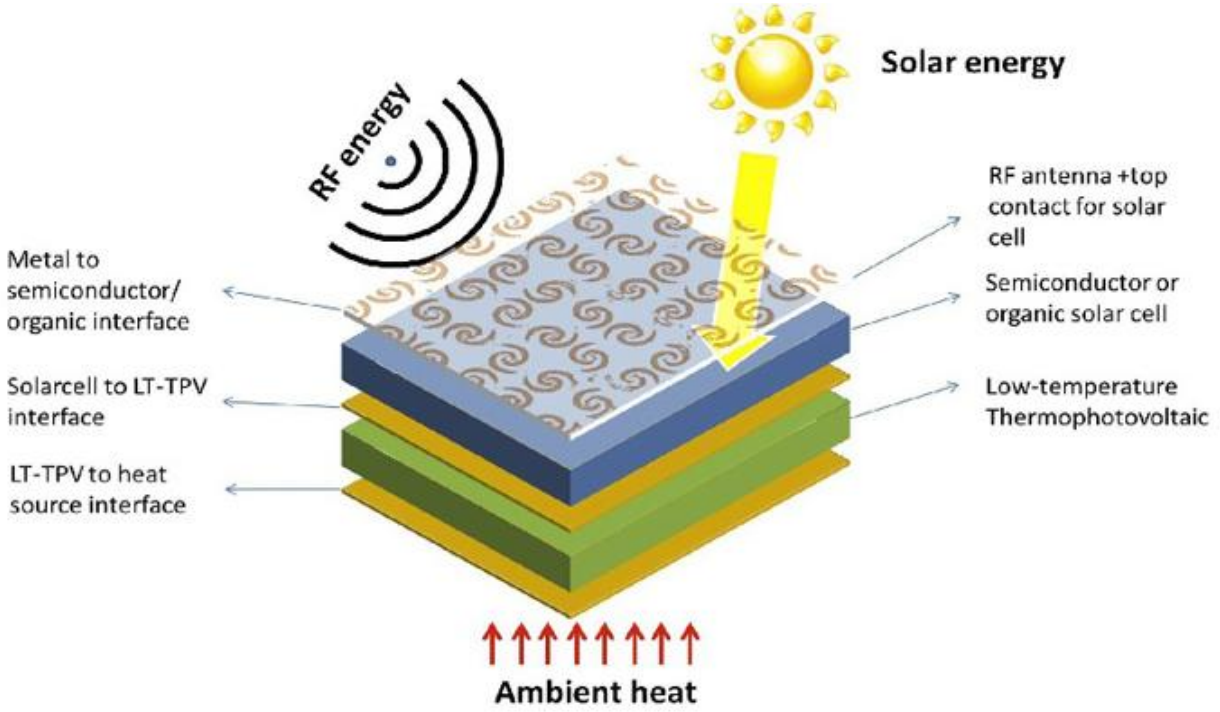
Thermophotovoltaic Harvesting

Such a system, therefore, combines the advantages of solar photovoltaic systems, which turn sunlight directly into electricity, and solar thermal systems, which can have an advantage for delayed use because heat can be more easily stored than electricity. The new solar thermophotovoltaic systems, they say, could provide efficiency because of their broadband absorption of sunlight; scalability and compactness, because they are based on existing chip-manufacturing technology; and ease of energy storage, because of their reliance on heat.





Future of Energy Harvesting



A monolithic energy harvester device, combining solar photovoltaic, thermophotovoltaic, RF and vibration energy harvesting.

Energy Harvesting Using a Cheap Easy-to-Fabricate FM Rectenna

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Abstract- **Energy harvesting techniques are emerging as environmental friendly energy sources, which form a promising alternative to existing energy resources. These include rectennas, solar cells, harvesting human energy and wind power. GHz/THz rectennas already exist and have some characteristics such as small size and wideband. These rectennas have a lot of disadvantages such as power fading, complicated design procedure and high fabrication technology. Thus, the present paper suggests using FM rectennas to solve these problems and by using fractals we will keep the advantages.**

The present paper discusses the design and implementation of a rectenna prototype operating in the FM range. Using Fractal antennas (which offer broadband/multiband operation) and simple tricks from circuit design, God's willing, proves to be a successful strategy for obtaining useful power from a cheap easy to fabricate rectenna that can be even made in home by interested users. The paper includes theoretical discussion, simulation results and practical prototype results.

Keywords: **rectennas, fractals, rectifying circuits, antennas, energy harvesting.**

I. INTRODUCTION

Energy harvesting techniques are emerging as environmental friendly energy sources, which form a promising alternative to existing energy resources. These include energy harvesting from rectennas, passive human power, wind energy and solar power.

Energy harvesting from human power is used to generate useful power by portable equipments are wearable devices in which digital systems are integrated in everyday personal belongings, like clothes, watch, glasses, etc. Human body can be considered as a storehouse of energy. There exist two possibilities: power can be scavenged from the user's everyday actions or can be intentionally generated by the user. This method has a lot of advantages. It allows the decrease in both size and power consumption of complex digital systems.

The power harvesting can be stored in capacitors, rechargeable batteries, etc. The disadvantages of this method arise because the portable products are powered by rechargeable batteries and they will remain as the main source for this kind of consumer products. However the disadvantage of batteries is the need to rather place or recharge them periodically. Typical candidate applications are extracting power from pedaling which can generate power up to 1.6 mJ, from typing which can generate pewer up to 2 mJ with every push of the button of 15 N and power from body heat which can recoverable 2.8-4.8 W of power.[1]

Another method to harvest energy is offshore wind energy where researchers are developing new technologies to provide electricity offshore. The new Spar-WARP wind machine developed by ENECO can produce and store clean, safe electricity offshore in many areas where there is sufficient wind speed. Most offshore areas provide good sites for wind technology. A good land site usually has mean wind speeds of 13 to 17 mph, while typical offshore mean wind speeds range from 15 to 20 mph [2].

The WARP, which stands for Wind Amplified Rotor Platform, is designed to further amplify wind by as much as 50 to 80 percent, and has many advantages for providing electricity in offshore areas. It can be installed in any depth of water—on a foundation in shallow water or on a floating platform tethered to the bottom by cable in deep water.

A third method to harvest energy the solar power may be used. Solar energy is an environment energy available to power portable devices. A photovoltaic system generates electricity by the conversion of the sun's energy into electricity.

Photovoltaic systems are found from the Megawatt to the milliwatt range producing electricity for a wide range of applications: from wristwatch to grid-connected PV systems.

The examples of consumer products that contain PV solar cells are: calculators, radios, headphones, laptops, battery chargers, etc. [1]

Moreover, we can use solar rectennas to harvest solar power but those rectennas are operating at GHz/THz and have a lot of disadvantages such as power fading, complicated design procedure and high fabrication technology. Thus, it has been suggested to use FM rectennas operating at FM range of frequencies using fractal antennas not to harvest solar power but to harvest electromagnetic waves instead of all those energy harvesting techniques because those FM rectennas are simpler and cheaper to construct. To maintain rectennas advantages, fractal antennas are used. DC-up converter circuits can be used to raise the voltage harvested by the rectenna.

The organization of the remaining of the paper is as follows. In section II, we will explain the rectenna concept and offer a brief literature survey which shows the efforts of some researchers. Section III explains fractal antennas and especially Sierpinski triangle fractals. This section will show the complete rectenna design applied of a fractal antenna called FRACTENT [5]. Finally in section IV the simulation and the practical results of complete rectenna are given including steps of constructing the cheap simple prototype. Simulations are performed using NEC2 software and P-SPICE software.

II. RECTENNA CONCEPT AND A BRIEF LITERATURE SURVEY

A rectenna is a device that can harvest microwave energy. It can be constructed using an antenna, rectifier circuit, smoothing circuit. A basic rectenna is shown in Fig.1. [1]

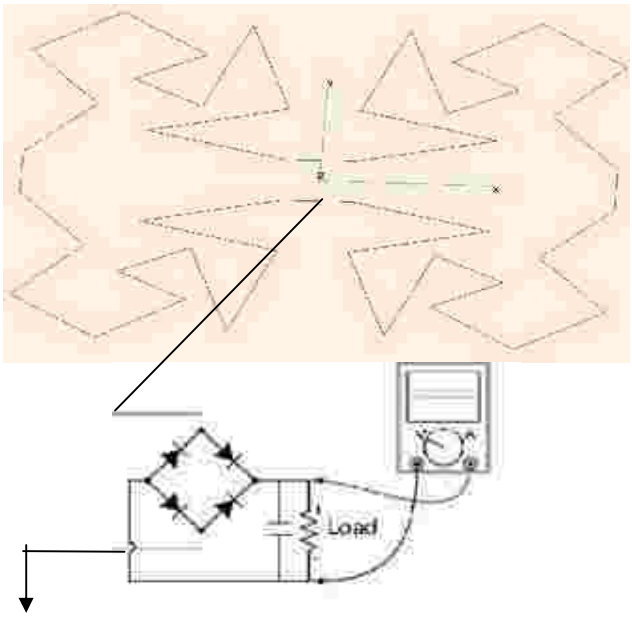

Fig.1 The complete rectenna

Several researchers reported the successful design and implementation of rectenna. J. Hagerty, ET. Al. have developed a new approach for construction of efficient rectenna arrays for arbitrarily polarized incident waves with broad spectral content. The approach has been validated experimentally on a dense grid array that rectifies two orthogonal linear polarizations, and on a self-similar spiral array with alternating right-hand and left-hand circular polarizations. The two arrays operate from 4.5to 8GHz and 8.5to 15GHz and have maximum open circuit voltages of 3.5 and 4.0V, respectively. Their efficiencies increase above 35% and 45%, respectively, for higher incident powers. The grid of rectifiers has excellent reliability and graceful degradation. The limiting factor on the size of the grid rectifier is the current rating on the diodes. Note that in the grid in Fig.2 the four corner diodes are the most critical ones, because one half of the current through the DC terminals passes through each of the diodes. If these 4 diodes are replaced by shorts, the current in the DC leads can be twice as large. The current intensity in the next diodes closest to the terminals is half of that through the four most critical diodes. Any overloaded diodes in the grid are automatically eliminated if they fail as shorts, and the rest of the grid continues to function. If a diode fails as an open, the current will find a path through the other diodes/shorts across the grid. [3]

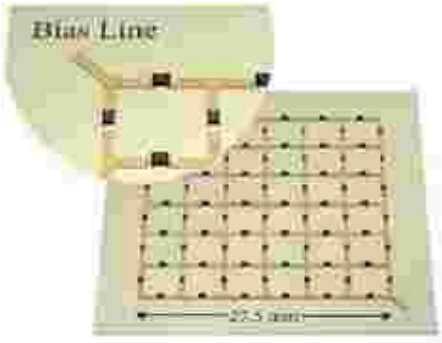

Fig.2 The grid array rectifier

Another researcher Y. Li has developed a 2.45 GHz low power rectenna design for wireless sensor and RFID applications. The system is monolithically integrated onto a single circuit board. A high gain patch antenna array is used to boost the power level at the input of the Rectenna for better power conversion efficiency. Moreover, a hybrid ring coupler is used to divide the power between Rectenna and detector. The system implements a key RF front end for GAP4S wireless sensor system [4].

III. FRACTAL ANTENNAS

Fractals are geometrical shapes, which are self-similar, repeating themselves at different scales. In this chapter we will shown two examples of fractal antenna (Sierpinski triangle antenna and kuch curve antenna), after that we will shown FRACTENT that will using in our FM rectenna.

The Sierpinski triangle, shown in Fig.3, is a common self-similar geometrical figure. It also has been used as a very effective antenna in the GHz frequency range.

The geometric construction of such a triangle is simple. One starts with the black equilateral shape and takes afterwards, in different steps, the middle of the sides and generates respectively 3, 9, 27, 81, triangles which are self similar and exactly scaled down versions of the initiating shape. The same procedure can be observed in Fig.4 where a Koch curve is iterated in 3 steps.

It is interesting to know something about the “Dimension” of such a fractured structure. The term “Dimension” in mathematics has different meanings. The common definition is the “Topologic Dimension” in which a point has the dimension 0, a line has the dimension 1, a surface has the dimension 2 and a cube has dimension 3.

Using Fractal geometry in antennas results in multi-frequency/broadband behavior. This behavior is best illustrated based on discussion of Sierpinski monopoles.

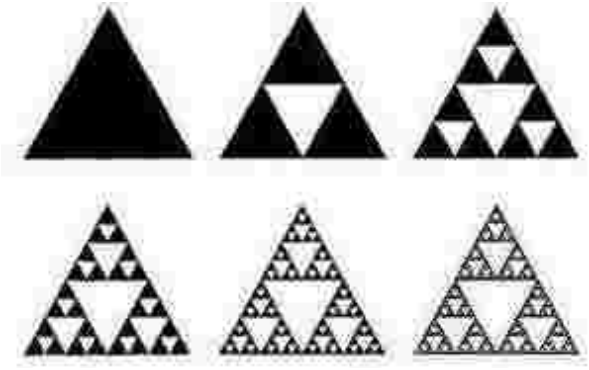

Fig.3 The Sierpinski triangle antenna

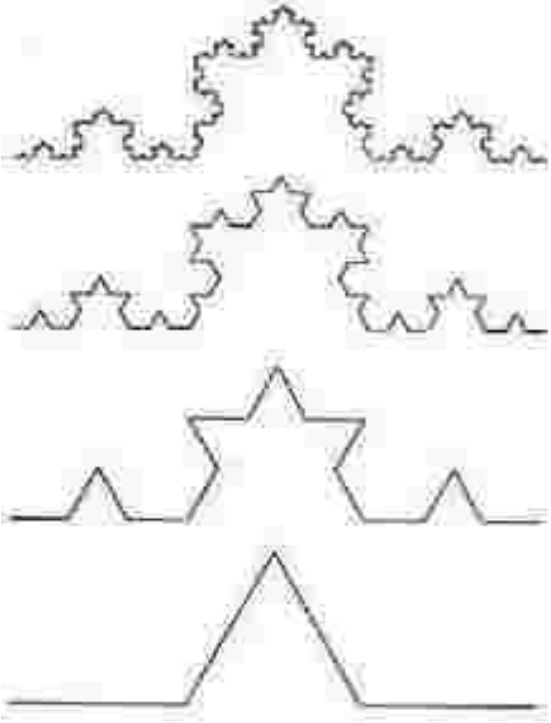

Fig.4 The Kuch curve antenna

The Sierpinski Monopole is shown in Fig.5. This is a monopole antenna resonant at frequencies of 0.44, 1.75, 3.51, 7.01 and 13.89GHz having an input resistance of 50Ω. One can easily see the 5 resonant frequencies of the structure by looking to the 5 circles marking the respective triangles.

The complete rectenna design (antenna+ rectifier and smoothing circuit) has been shown in Fig.1.

The multi broadband response of fractal antennas mentioned earlier make them suitable for use in rectenna applications since they can harvest more power than a single band antenna. The FRACTENT [5] is a particularly interesting antenna geometry as shown in Fig.6.

Since the Fractent operates at FM range, the author was able to construct it using the following simple procedure. First, the Fractent operating was drawn to scale. Second, a common 1-mm copper wire was shaped by hand to take the Fractent shape.

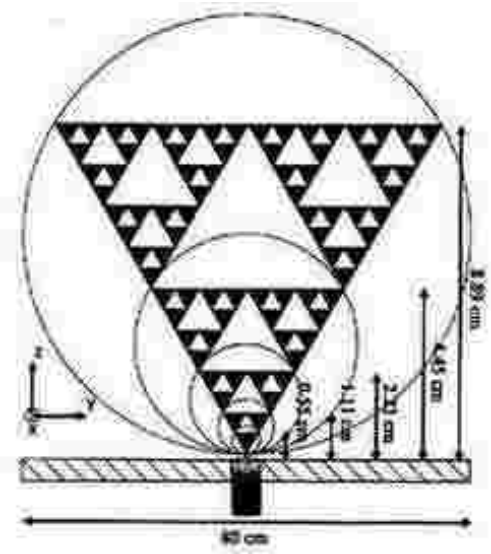

Fig.5 Sierpinski Monopole antenna resonant

Fig.6 Fractent antenna

The performance of FRACTENT has been studied through simulation using NEC2 software and has been validated practically by the author by putting it into operation as an FM antenna. The rectification circuit and smoothing circuit shown in Fig.1

The following design formula has been used Full-wave bridge rectifier with Schottky diode. Schottky diodes have been used duo to their low turn-on voltage and fast response which is necessary at high frequency.

It is important to note that the choice of FM range is important because it allows for easy and cheap rectifier/smoother circuit design and implementation while working in higher microwave frequency range requires more technologically complex methods.

IV. SIMULATION AND PRACTICAL RESULTS

The fractal antenna prototype constructed by the author is shown in Fig.7

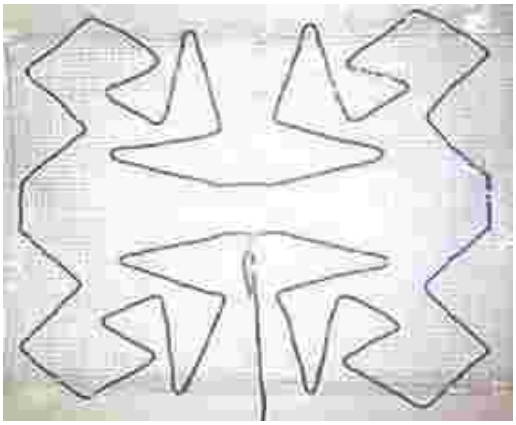

Fig.7 The cheap practical prototype made by the author

We use NEC2 software to simulate this antenna and results at 90 MHz are shown in Fig.8.The antenna radiation pattern is shown in Fig.9. Simple frequency scaling was used to change the dimensions of the antenna so as to transform its bandwidth to 90 MHz.

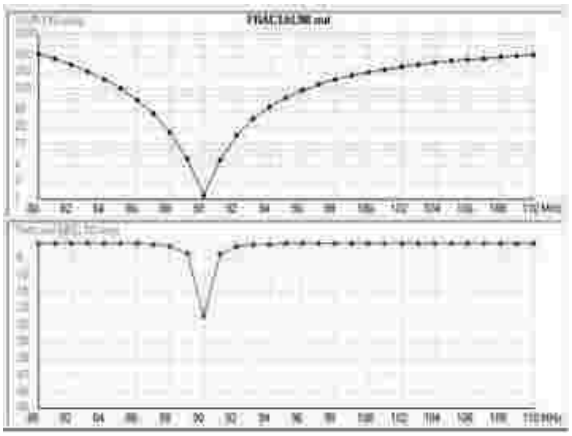

Fig.8 NEC2 simulation results for the practical prototype (VSWR and reflection coefficient in dB)

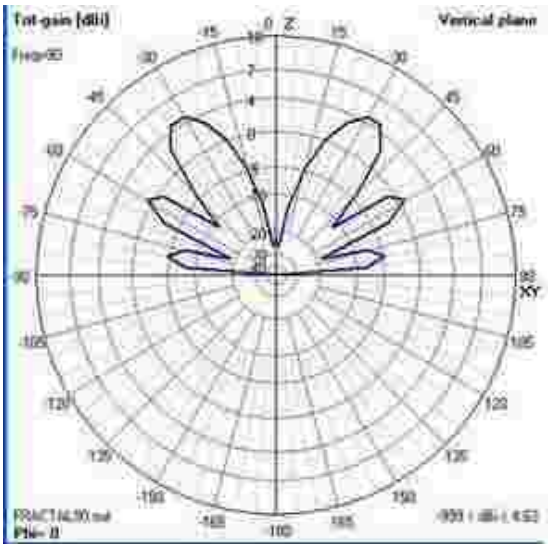

Fig.9 The Fractal antenna radiation pattern in the horizontal plane at 90 MHz

Rectifier circuit simulate on P-SPICE program. The results of P-SPICE program are shown in Fig.10.

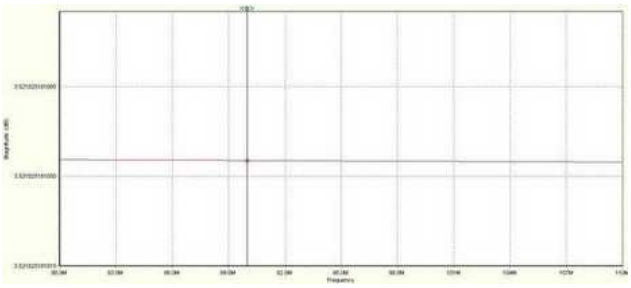

Fig.10 P-SPICE program results for usage rectifier circuit

V. CONCLUSIONS

In this paper, it was shown that energy can be harvested from some sources such as rectennas, human power, solar energy and offshore wind energy. The rectenna concept was the focus of the paper as efficient alternative to common GHz/THz rectennas. To get over the technological difficulties associated with developing rectennas at GHz/THz range, the author developed a new rectenna efficient design that uses fractal antennas operating at FM range. The rectenna operating an FM range is easily constructed using common copper wire, diodes and capacitors, which make them easy and cheap to develop even by inexperienced users without compromising rectenna efficiency. Thus, the developed rectenna is cheap, efficient, simple to implement and environmental friendly.

ACKNOWLEDGEMENT

Thanks to Allah almighty for helping us complete this work. Thanks to my colleagues for their support.

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The Rectenna Device - from theory to practice (a review)

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The Rectenna (RECTifying antENNA), which was first demonstrated by William C. Brown in 1964 as a receiver for microwave power transmission, is now increasingly researched as a means of harvesting solar radiation. Tapping into the growing photovoltaic market, the attraction of the rectenna concept is the potential for devices that, in theory, are not limited in efficiency by the Shockley-Queisser limit. In this review, the history and operation of this 40-year old device concept is explored in the context of power transmission and the ever increasing interest in its potential applications at THz frequencies, through the infra-red and visible spectra. Recent modelling approaches that have predicted controversially high efficiency values at these frequencies are critically examined. It is proposed that to unlock any of the promised potential in the solar rectenna concept, there is a need for each constituent part to be improved beyond the current best performance, with the existing nanometer scale antennas, the rectification and the impedance matching solutions all falling short of the necessary efficiencies at THz frequencies. Advances in the fabrication, characterisation and understanding of the antenna and the rectifier are reviewed, and common solar rectenna design approaches are summarised. Finally, the socio-economic impact of success in this field is discussed and future work is proposed.

I. INTRODUCTION

The ever increasing global demand for clean renewable energy is a crucial topic that has major economic and societal impact for our future on this planet. The sun is a constant source of energy, thus better harvesting its radiation would solve the energy challenge and help protect the environment. Conventional solar harvesting is via the increasingly popular photovoltaic (PV) p-n junction solar cells. Although this field has seen great improvement over the last few decades since the first silicon solar cell was developed in 1941 at Bell Laboratories, its fundamental limitations discussed by Shockley and Queisser[1] in 1961 puts an upper boundary on the efficiency at around 30% for single junction solar cells[1] and 55% for multi-junction solar cells[2]. Due to the efficiency limits of solar cells, other solar-energy converting technologies are becoming increasingly attractive. An alternative solar energy converter concept is the Rectenna (RECTifying antENNA) device, which has achieved very high efficiencies at microwave frequencies, while at THz frequencies, it has been speculated that efficiencies exceeding the Shockley-Queisser limit are achievable.

Fundamentally the Rectenna and the Solar Cell are similar as there is absorption of photons to generate a DC current. However, there is a substantial difference in their principles of operation. Solar cells are quantum devices, except for their anti-reflection coatings, and are only able to be understood and designed by the application of quantum physics[3], where light is perceived in its particle form, as photons. Hence, the efficiency is limited by the band gap energy of the active layers, where energy is lost due to low energy photons passing right through the material, and through heat loss from the high frequency photons. The rectenna device concept is different as it is based upon the wave nature of light, which is regarded as a propagating, oscillating electric and magnetic field[4]. Similar to a low pass filter in electronics, the efficient absorption of solar radiation is limited by the cut-off frequency of the device, i.e. any wavelength below this frequency can be harvested.

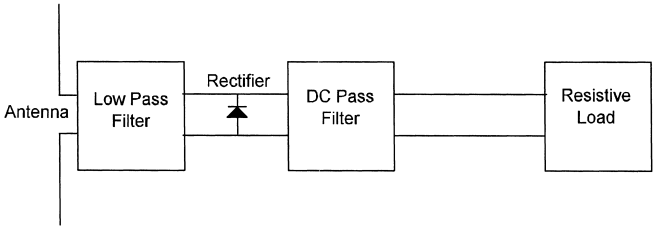

FIG. 1. Block diagram of microwave rectenna and load. (Reprinted from [3], ©2002 with permission from Elsevier)

The components of the rectenna device are shown in Figure 1. These components are commonly associated with a rectenna operating at microwave frequencies. The antenna is used to capture and convert the energy from an incoming microwave beam (or solar radiation in the case of solar rectenna) as it is an efficient transducer between free space and guided waves[5]. The electric field (**E**) from an incident electromagnetic wave induces an

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alternating current (AC) in the conducting antenna. A low pass filter is used to form an impedance match between the antenna and the rectifier, and passes the signal at the tuned operating frequency whilst blocking reradiation of higher order harmonic energy produced by the rectifier[3,6–8]. The rectifier (a diode) converts the input AC signal to a usable direct current (DC) which is transferred to the load. A DC pass filter, consisting of an inductor in series, is used to smooth the rectified signal to DC whilst providing efficient transfer to the load.

The antenna dimensions are of the same length scale as the targeted frequency[5]. Therefore, for microwave frequencies, the antenna is relatively large (cm to mm scale) compared to that required for infrared (µm scale) and optical frequencies (nm scale). Due to its smaller size, it will be difficult to incorporate the aforementioned filters into the design of a rectenna for solar energy collection. Therefore the components of the solar rectenna are reduced to just the Antenna and Rectifier (Figure 2)[9]. The removal of the low-pass filter from this design enables broadband frequency harvesting up to the cut-off frequency of the device, although it introduces the difficulty of impedance matching the antenna and rectifier, resulting in a potential drop in efficiency. Losses in transferring the rectified DC signal to the load are dependent on the diode behaviour. However, appropriate design of a DC pass filter would make its use possible in this configuration[99,100].

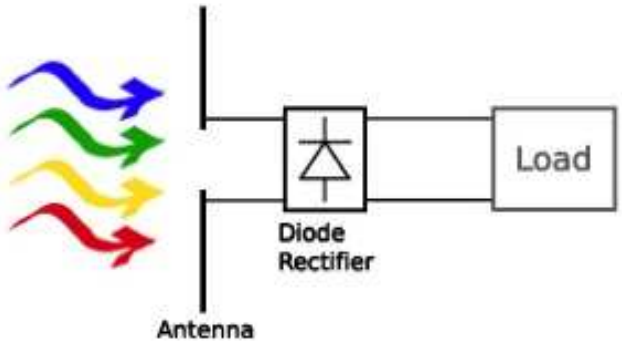

FIG. 2. Block diagram of solar rectenna and load. (Reprinted from [9], ©IOP Publishing. Reproduced by permission of IOP Publishing. All rights reserved.)

There are two families of designs for rectenna solar energy collectors. Bailey[10] proposed that each antenna element is attached to its individual rectifier and the DC outputs are combined (see Figure 5). The immediate concern here is that the small voltage predicted from the antenna elements may be insufficient for rectification to occur in the attached diode element[3]. The only way to overcome this is to design a diode which turns on at a voltage as low as that arising from the single antenna element. The other design is by Kraus[11], who suggested that the electrical oscillations from many antenna elements be combined in a particular phase relationship, then delivered together to the rectifier. Limitations in this design arise from the need for spatial coherence across all the antenna elements feeding the signal to the diode[3]. This can be minimised by appropriate design and high quality fabrication of the antenna elements.

The aim of this article is to review key research in the development of the rectenna device as an alternative or in addition to conventional solar cell technology. We aim to expand upon a few short reviews, [3–5,9,100,105], that have been published, bringing the reader up to date with a more detailed overview of the device. This article will be the first to bring together all relevant research not only on the rectenna as a whole device, but also the research on its two main components – the antenna and the rectifier. The next section will briefly describe the history of the device and how the initial application of microwave power transmission (MPT) was later modified to a solar energy harvesting alternative. Section III will analyse and discuss the many publications on experimentally and theoretically derived device efficiencies. Breakthroughs in modelling and fabricating the antenna and rectifier will be reviewed in Section IV and V respectively. Section VI will bring together all components and look at different rectenna architectures, which show potential for efficient operation at solar frequencies. We will discuss the socio-economic impact of the research in this field and proposed future work in Section VII before completing the article with a short summary in Section VIII.

II. HISTORY OF THE RECTENNA DEVICE

The rectenna concept was initially proposed for wireless power transmission by William C. Brown of the Raytheon Company in the 1960's. At the time, Brown, a pioneer of microwave power transmission, was faced with a challenge of how to efficiently receive and convert to DC power an incoming microwave beam. Wireless power transmission was not a new concept as it had begun with the ideas and demonstrations by Tesla in 1899 using radio waves[12]. Tesla managed to demonstrate this concept although never found an applicable route into its commercialisation.

In 1959 the Raytheon Company proposed the Raytheon Airborne Microwave Platform (RAMP) concept to the US Department of Defence as a solution to communications and surveillance problems[13]. This proposed platform was to be flown at high altitude (50,000 ft) in the form of a helicopter, powered by microwave transmission. The US Department of Defence did not fund the project however the Raytheon Company pursued their interest by researching the required technological developments for its realization. One missing technology was a suitable microwave rectifier[14], capable of converting the microwave signal directly into the DC current necessary to power the platform's motors and stay airborne. Their initial design used a

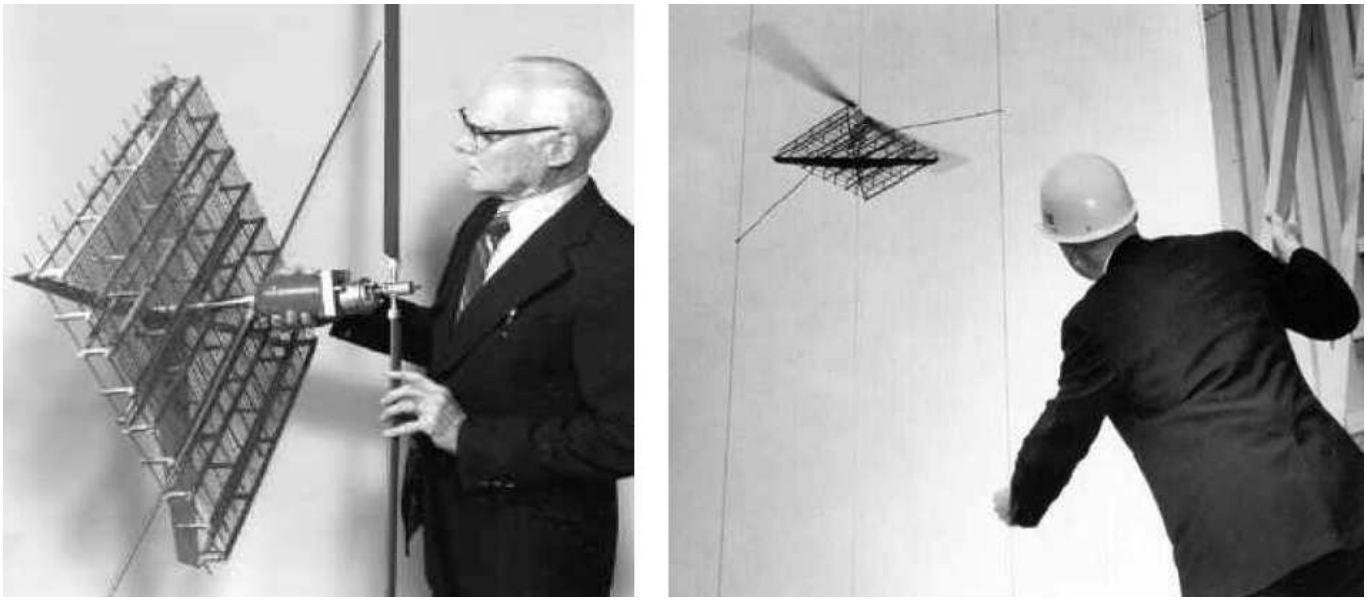

FIG. 3. Left: W.C. Brown holding the rectenna helicopter platform. Right: In flight demonstration of the rectenna helicopter on October 1964. (Reprinted from [17])

horn-illuminated ellipsoidal reflector that focused the microwave beam into a trapezoidal horn at a 20 ft distance where the microwaves were converted to DC power by the rectifier[13]. Using this first design, Brown constructed and flew thirty feet above his lawn a small tethered helicopter that he had built with his son in his own workshop[13]. Although this concept worked, Brown was aware of the flaws in his system and continued to work on other possibilities.

In 1963, Brown met with his colleague and friend Roscoe George, a professor at Purdue University who had been working on a microwave point-contact semiconductor diode[15]. Following discussions regarding his challenge, a device which is now known as "Rectenna" was conceived. The proposed solution was patented in 1969 and consisted of taking individual full-wave rectifiers out of the waveguide, attaching them to half-wave dipoles, and placing a reflecting plate behind the structure[16]. The first microwave rectenna was built by R.H. George at Purdue University but conceived at Raytheon Company in 1963, shown in Figure 4. Later in 1964, Brown built a helicopter which had an array of 28 such Rectennas. The helicopter was demonstrated in flight 60 ft above a transmitting microwave beam on the CBS Walter Cronkite News in October 1964. Figure 3 shows some impressive images of this historic event. The interested reader is referred to Brown's recollection on "The History of Power Transmission by Radio Waves" in [13], where all the interesting details and facts about the invention of this device can be found.

Following this invention, microwave power transmission became very popular and many potential applications were researched. An interesting one is the solar

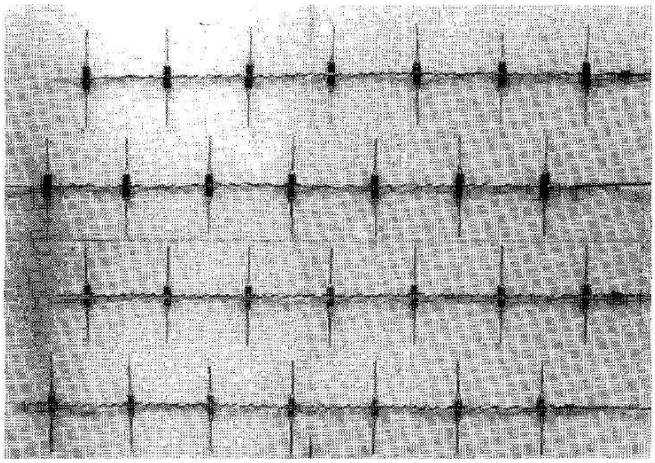

FIG. 4. The first rectenna. Conceived at Raytheon Company in 1963, built and tested by R.H. George at Purdue University. It was composed of 28 half–wave dipoles, each terminated in a bridge rectifier made from four 1N82G point–contact, semiconductor diodes for operation at 2.45 GHz. A power output of 7W was produced at 40% efficiency. (Reprinted from [13], ©1984 IEEE. Reprinted, with permission, from IEEE)

power satellite (SPS), proposed by P. E. Glaser[18] in 1968. The concept of the SPS represents a 5 GW power station on the geostationary orbit, which collects electrical power by means of solar cells and transmits microwaves of the 2.45 GHz band to a rectenna array site on the ground. The SPS project was researched by Brown and Glaser[19–22] which earned Glaser a patent in 1973[23] and

a collaborative three year project together with Brown in 1977 called the "DOE/NASA Satellite Power System Concept Development and Evaluation Program". The work ended in 1980 with much success in developing new ideas in the design of transmission technology and rectenna arrays however did not attract further funding. Although this was the end of the SPS initiative in the US, some experimental studies of the concept were seen in the 1990s in Yamasaki, Japan[24,25]. The interested reader is referred to Glaser's overview of SPS progress in [26].

The developments from the SPS project yielded other applications involving microwave transmission and rectenna, such as intersatellite power transmission[27], including utility power satellites[28], mechanical actuators for space-based telescopes[29], small dc motor driving[30], short range power transfer as for example between two parts of a satellite, RF identification tags (RFID) and biomedical implants[31]. These applications all include single band rectenna operation. Hagerty *et al*[31] have shown a broadband rectenna array for microwave operation to be used in low-power indoor sensor networks and RF energy recycling. Recent publications have shown a compact printed rectenna that can be used to supply DC power to electrical devices supporting data communication systems[32].

The idea of exploiting the wave nature of light, or solar radiation, and converting its energy into usable direct current by the use of antennas, originated as part of a NASA summer project in the late 1960s undertaken by J.C. Fletcher of NASA and R.L. Bailey of the University of Florida[4]. The first official report of this concept was published by Bailey in 1972[10]. In this paper, he named his device the "*Electromagnetic Wave Energy Converter (EWEC)*" and suggested an artificial pyramid or cone structure as antennas, analogous to those found in nature and similar to dielectric rod antennas. The paper described pairs of pyramids as modified dipole antennas, each pair electrically connected to a half-wave rectifying diode, filter and load. This invention is fundamentally very similar to the initial microwave rectenna proposed by Brown, but it is modified so that the antenna and rectifier operate at much high frequencies to harvest solar energy. The device called by Bailey EWEC is now referred to as a Solar Rectenna.

Bailey's first publication on the subject in 1972 served as a proof of concept as he showed the broadband characteristics of his pyramidal antennas by tests at 100-1000 MHz. Following this, in 1973, Fletcher and Bailey published a patent[33] on the EWEC concept. Later in 1975, in their final report, Bailey *et al*[34] presented their theoretical work highlighting the importance of future research into optimised antenna design and fabrication, while not forgetting that rectification of the signals was an equally critical issue and a significant challenge[4]. These two components of the rectenna solar energy harvester are now of equal research interest (see Sections IV & V).

Later, alternative rectenna structures were proposed

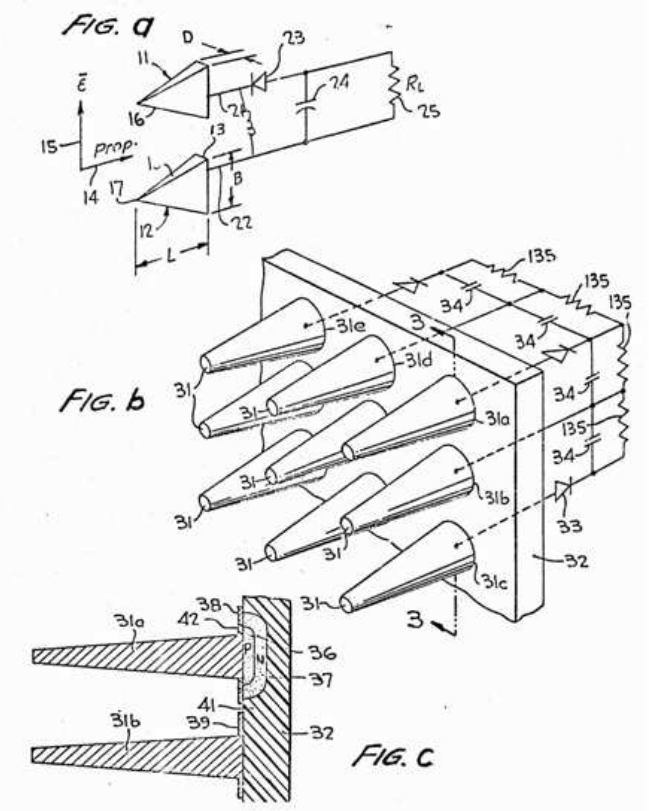

FIG. 5. The EWEC concept by Fletcher & Bailey. Each broadband conical antenna is connected to a rectifying diode, filters and load. (Reprinted from [33])

and patented by Marks[35–38], who was inspired by the earlier half-wave rectifier EWEC concept. His first patent in 1984 was a device with arrays of submicron crossed dipoles on an insulating sheet with fast full-wave rectifiers[35]. Later in 1988, Marks was first to patent antenna-like cylinders attached to asymmetrical metal-insulator-metal (MIM) diodes for rectification[37]. MIM diodes are one of the most promising rectifiers for use in Infrared and Optical Rectennas and will be later discussed (Section V.B).

The first reported experimental evidence of a fabricated resonant nanostructure absorbing and rectifying at visible light frequency was published by Lin *et al*[39] in 1996. They recorded a short circuit current using a parallel dipole sub-nanostructures connected to a p-n junction for the rectification of the absorbed signal. Credit must be given also to Gustafson & Billman[40] who first suggested harvesting visible radiation in 1974. Their work formed the basis of using optical diodes like the MIM diode.

Figure 6 outlines a brief timeline of the key events that have lead to the current research activity in solar rectenna. Further reading is suggested in [3,4] for a more detailed reference and explanation of the major historic breakthroughs in the field of solar harvesting by rectenna.

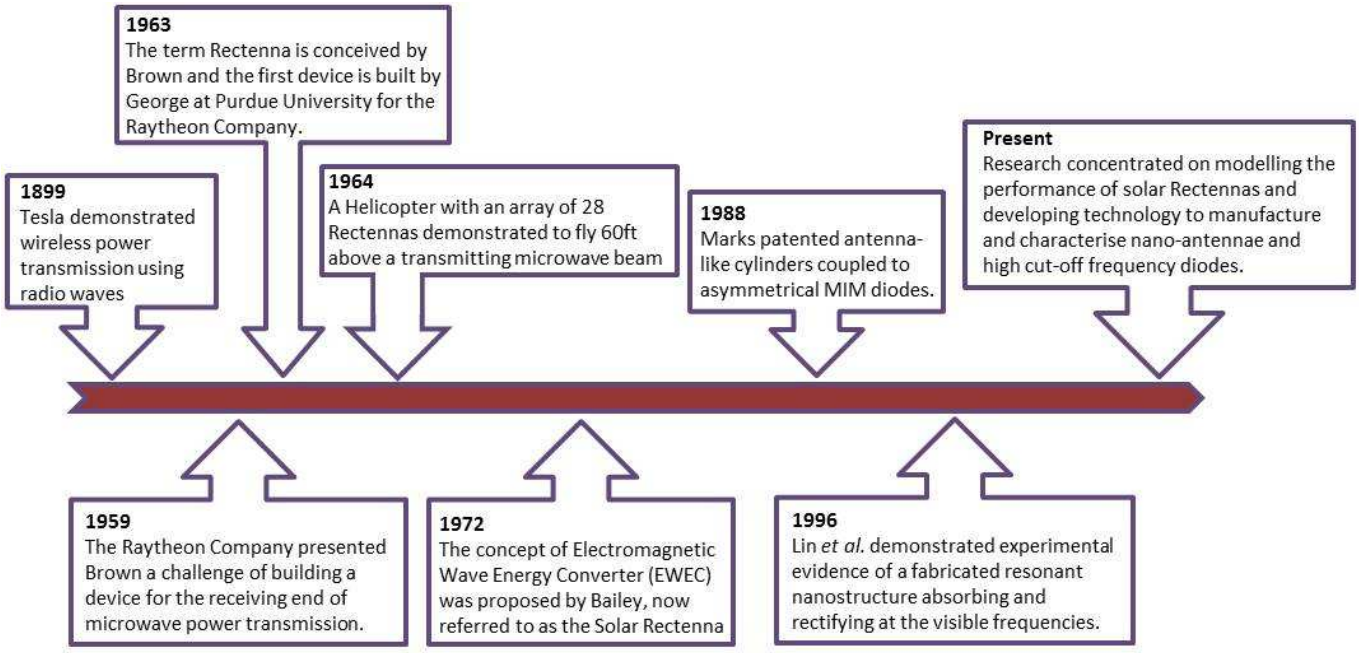

FIG. 6. Brief timeline of the rectenna device - from concept for microwave power transmission to research focused on solar harvesting

In the following sections, we mainly focus on the key research in harvesting optical and infrared frequencies using the rectenna concept.

III. EFFICIENCY LIMITS

The first rectenna device (Figure 4), which was built in 1963, provided a 7 W power output with 40% efficiency[13] designed to work at 2.45 GHz. This was the most commonly targeted transmitting frequency for MPT because of its low attenuation through the atmosphere even in severe weather and being at the centre of the industrial, scientific, and medical (ISM) band meant that it was the most advanced and efficient transmitting technology.

This rectenna used half-wave dipole antennas to receive the signal and point-contact semiconductor diodes for rectification. In later device attempts, the half-wave dipole antenna remained the preferred technology, but an increase in efficiency was mainly achieved by improving the rectifier performance, incorporating Schottky Barrier Diodes (SBD). From the early 1960's until the early 1980's there has been a steady increase in the experimentally demonstrated efficiency of the rectenna at 2.45 GHz (Figure 7a). Although Nahas[7] first developed a model (using a distributed transmission-line technique including skin-effect losses) to simulate the top performance of a rectenna utilising a dipole antenna and a conventional SBD, his prediction of an 80% upper boundary (most of the losses came from the diode rectifier) was later improved when Brown[82] experimentally demonstrated what is to date, the highest ever recorded efficiency at the 2.45 GHz band of 92%. This was made possible by using the then well established Pt/GaAs Schottky diode and supported Brown's predictions that 100% efficiency is possible for this device[44,82].

In another attempt to model the efficiency, Gutman & Borrego[51] predicted an upper boundary of 85% incorporating numerous rectifier outputs interconnected to a common DC load. They used a closed-form analytical circuit model and a computer-simulation model in an attempt to show that a higher output power can be achieved having an array of these components as opposed to a single rectenna element.

It is important to state that the conversion efficiencies (η) considered here are simply defined as the ratio of the output power (P_{out}) over the power incident on the antenna (P_{in}),

$$\eta(\%) = \frac{P_{out}}{P_{in}} \times 100 \tag{1}$$

One of the most successful demonstrations of MPT was in 1975 by Dickinson and Brown[42] who used an array of 5000 rectennas to receive a transmitted signal over a distance of one mile with 82% efficiency at 2.388 GHz[42,43]. The rectenna architecture again contained half-wave dipole antennas. In the last two decades many other antenna architectures have been proposed in search of enhanced performance and economical alternative. These include the microstrip patch[44,46], circular polarised[53,56–58], compact slot ring[55]

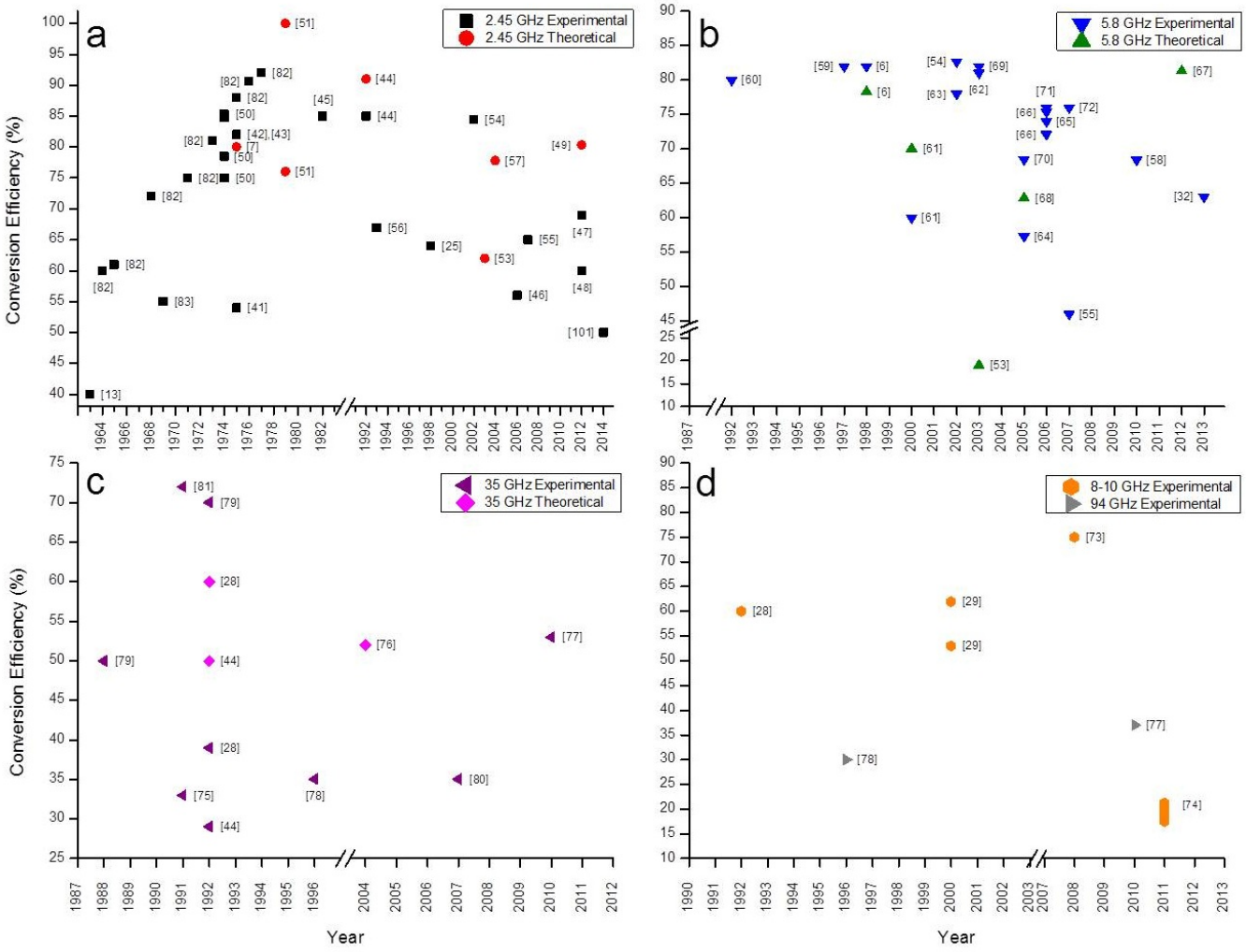

FIG. 7. Reported conversion efficiency (as defined by their authors) evolution over the years of a rectenna device for Microwave Power Transmission. Data points are split into Experimental and Theoretically derived values of efficiency. a) 2.45 GHz, b) 5.8 GHz, c) 35 GHz and d) 8-10 GHz and 94 GHz. [**] indicates the reference corresponding to the data point. Data collected from [3,5–7,10,11,13,25,28,29,32,41–51,53–82,101]

and shorted annular ring-slot[47,53] antennas with all attempts returning efficiencies between 55 and 85% (see Figure 7a). The higher efficiencies reported all used GaAs Schottky diodes as opposed to conventional Si Schottky diodes, showing the importance of diode behaviour on device performance.

In the 1990's a renewed interest in the solar power satellite (SPS) concept[18–26] started a new focus of research into high efficiency rectennas and rectenna arrays operating at higher frequency than the mid-ISM band (2.45 GHz)[41]. With higher operating frequency, the device dimensions can be reduced conveniently from the point of SPS construction feasibility[41]. Also small size design can be appropriately implemented in a range of other applications such as a supply source in RFID[31,55,58,64,68,77], wireless sensors[31,47,58,77] and data communication systems[32,61,64,65,67,77]. Rectennas operating in the C-band (≈ 5.8 GHz – Figure 7b), X-band (8-10 GHz Figure 7d), Kα-band (35 GHz Figure 7c) and W-band (94 GHz Figure 7d) have appeared after the 1990s, some even operating in two bands (2.45 & 5.8 GHz[53–55] and 35 & 94 GHz[77]). However, the efficiency is different in the two frequency bands with the higher frequency operation yielding lower efficiency due to losses in transferring the AC signal generated from the antenna to the rectifier circuit. The only dual frequency rectenna with high efficiency in both bands is reported by Suh & Chang[54] with an excess of 82% efficiency at both 2.45 and 5.8 GHz employing a CPS dipole antenna and GaAs SBD.

At 5.8 GHz, efficiency values in the range of 60-85% have been both predicted[6,61,67,68] and demonstrated[6,32,54,59–63,65,66,69–72] with different approaches returning equally successful devices through

the years (Figure 7b). The few experimentally reported rectennas at 8-10 GHz have seen slight improvement in efficiency when a circular or dual polarised printed patch antenna has been coupled to the GaAs[28,29] or Si[73] Schottky diode, however with a planar slot antenna[74] the efficiency is only 21.2% (Figure 7d). At 35 GHz (see Figure 7c) theoretical predictions are higher than their corresponding experimental results[28,44], although with a similar antenna and diode design, high experimental efficiency of 70% has been reported elsewhere [79,81]. Analysis of the different efficiencies achieved with similar rectenna technologies highlight the importance in the circuit and device design for rectennas used in microwave power transmission. Furthermore, load impedance and the power of the incoming beam also affect the overall efficiency of the device.

It can be noticed from Figure 7 that for a given frequency some models predict a lower efficiency than that which had already been experimentally achieved. The reason is that over the years, new antenna and rectifier technologies, and alternative, more compact circuit designs were emerging, which could not improve or even match the record efficiency achieved with the half-wave dipole antenna and Pt/GaAs SBD. Furthermore, scaling down the components and aiming towards higher frequency operation at lower cost, introduced difficulties for appropriate circuit design and the efficiency suffered a continuous drop. Overall, considering the evolution of reported efficiencies over the years (Figure 7) for MPT rectennas, it can be concluded that with appropriate antenna design, good matching to a suitable well-functioning diode rectifier connected to a DC load, experimental evidence of efficiencies over the 70% mark can be obtained for frequencies up to 35 GHz (see also Figure 8).

The high efficiencies achieved at the low GHz frequencies are due to the ease of fabricating the devices and designing appropriate filters (low pass and DC pass – see Figure 1). To be able to receive a signal, the antenna has to be of the same length scale as the wavelength of the incoming wave, which at the low GHz frequencies ranges from a few mm to 10's of cm. At infrared and optical frequencies, wavelengths are from μm to nm, which subsequently indicates the necessity of building the antenna and rectifier in this length scale range. At this scale, a direct impedance match of the antenna to rectifier has to be achieved (Figure 2) instead of using a low-pass filter, which cannot be designed at such small dimensions. This is what Bailey's proposed electromagnetic wave energy converter[10] implies – a scaled down version of Brown's rectenna consisting of a μm length antenna connected directly to the diode rectifier transferring the output to the load through a DC pass filter of some sort. Bailey predicted an efficiency in excess of 50% for his EWEC device at microwave frequencies, although not backed by any modelling or experimental evidence of a whole device. The only evidence he gives is the ability of an EWEC prototype to absorb electromagnetic plane polarized waves at 475 MHz. His vision was to study the concept at microwaves and later scaling it to light frequencies where he speculates potential for unity efficiency. A MPT rectenna device without a low-pass filter is possible and was demonstrated in 2012 at 2.45 GHz[47]. The proposed printed shorted annular ring-slot rectenna is designed to match the impedance of the SBD rectifier at 2.45 GHz. The diode however will produce second and third order harmonics which will re-radiate at 4.9 and 7.35 GHz but these will be blocked because the antenna is designed to mismatch its impedance at the higher order harmonics thus preventing re-radiation[47]. This device architecture removes the need for a low-pass filter design and achieves 69% efficiency.

Apart from Brown[45,82] (for microwave) and Bailey[10] (for solar) several other authors (Kraus[11], Balanis[86], and Andersen & Frandsen[107]) have speculated about the possibility of 100% conversion efficiency coming from highly absorptive no-loss antennas and ideal matching, transferring and rectification of the absorbed input signal to the load. This antenna efficiency is assuming uniform distribution of energy across the antenna, a single frequency and a coherent source. The high efficiency values for MPT are achieved not only with appropriate antenna design, matching circuit and rectifier performance, but also with efficient delivery of a good signal, because the artificially created microwave signal is single band, polarised and coherent. This makes the antenna engineering easier. Simply reducing the size of the components to operate at solar frequencies does not necessary mean that the efficiency will remain as high as that observed for MPT Rectennas. Unlike monochromatic microwave radiation, sunlight is unpolarised, incoherent and distributed over a wide band of wavelengths[96]. This creates further challenges in antenna engineering and lead to disputes on how to model the device performance. Furthermore, the diode has to be optimised to rectify the signal at the higher frequencies and both antenna and diode have to be well matched to maximise power transfer.

Provisional tests at 10 GHz with a dipole antenna and a Schottky diode were demonstrated by ITN[5] estimating efficiencies of over 50% however this frequency is too low for solar energy harvesting. ITN state that the lower frequency rectenna is scalable to the higher THz frequencies required for solar harvesting with their models predicting up to 85%[5] efficiency. According to microwave antenna theory, the antenna length scales linearly with the incident frequency which means that in theory the antenna can be scaled to resonate at IR and optical frequencies[5]. This is not entirely the case because at IR and optical frequencies the conduction is not Ohmic and the scaling laws are rather an estimation than an accurate representation of the real-life situation as the majority of the energy in the surface modes is carried in the dielectric above the antenna. This symbolises the importance of impedance losses. The antenna has to be designed to resonate against a complex waveform

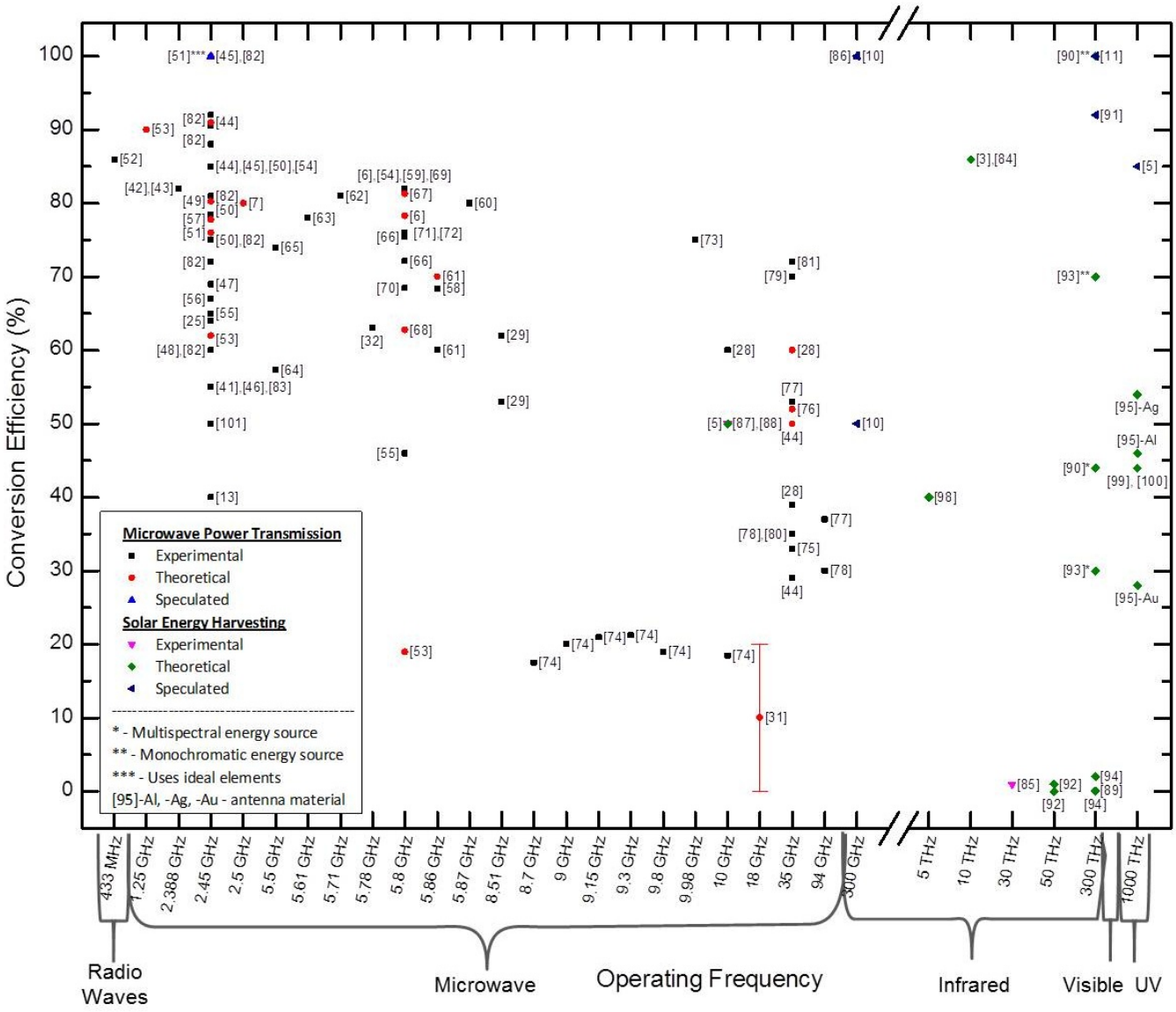

FIG. 8. Reported conversion efficiencies (as defined by their authors) at their maximum operating frequency of a rectenna device. Data points are grouped by the two applications of the rectenna - either Microwave Power Transmission or Solar Energy Harvesting, each displaying experimental, theoretical or speculated efficiency values. [**] indicates the reference corresponding to the data point. Data collected from [3,5–8,10,11,13,25,28,29,31,32,41–101]

and absorb broadband solar radiation which ranges from about 150 to 1000 THz, 85% of which corresponds to a wavelength of between 0.4μm to 1.6μm[5]. To reach the targeted frequencies in the high THz region, SBD rectifiers would not meet the criteria as the best cut-off frequencies are only up to a few terahertz[103]. MIM diodes are used for their potential to work in the higher THz region. The first practical demonstration was done by NIST[85] who used atomic layer deposition grown $Cr/CrO_x/Au$ diodes coupled to a dipole antenna and measured its efficiency at 30 THz (achieved using a CO_2 laser) to be below 1%. This attempt was carried out on an un-optimised antenna structure with a poor MIM diode. Following this, numerous theoretical models have been proposed, predicting the overall conversion efficiency of the solar rectenna (see Figure 8). For MPT rectennas, most theoretical and experimental efficiency results tend to be of the same order, whereas it can be seen that for Solar rectennas different attempts on modelling the performance predict a spread of efficiencies. Some approaches are close to the Landsberg limit of 93.3%[102] whilst others predict discouraging figures below 1% (Figure 8).

Conversion efficiencies from 0.001% to 2% have been

theoretically derived[89,92,94] employing various assumptions and a variation of modelling approaches (Figure 8). Mashaal & Gordon[89] analysed the efficiency in terms of losses by the state-of-art rectenna designs which adopt single polarization antennas able to convert only half of the random-polarization input power from solar radiation and a half-wave rectifier blocking the negative voltages of the input signal. Their assumptions were based on ideal matching between antenna and rectifier, zero turn-on voltage, limitless switching speeds and non-filtered signal output from the antenna prior to entering the rectifier which yielded an overall conversion efficiency of 0.0795% and four times as high if a full-wave rectifier is employed. It is a common misperception that with rectennas, the diode forms a half-wave rectifier, which would hold true for a single diode in a series configuration. However, common rectenna design incorporates a diode connected in parallel, which forms a clamp circuit capable of rectifying the full wave[99,100]. Therefore, estimating rectenna conversion efficiency should not be discouraged by assuming a half-wave rectification mechanism. Instead, efforts should be concentrated in understanding the effect of impedance mismatch. Briones *et al.*[92] based their model on the assumption that the collection of EM radiation by the antenna is done efficiently, which resulted in 0.001% efficiency of the system due to impedance mismatch losses between antenna and rectifier and inefficient rectification of the signal. By assuming efficient rectifier performance, the potential efficiency rises to approximately 1%.

An equally important criterion for efficient device operation is antenna material selection. Vandenbosch & Ma divided the analysis on the efficiency of nano-rectennas into two parts – 1) the efficiency by which the light is captured by the antenna and brought to its terminals[97] and 2) the efficiency by which the captured light is transformed to low frequency power by the diode[95,96]. Considering first the antenna efficiency, they studied various antenna materials in the form of 250 nm dipoles placed on a substrate[97]. Their calculations predicted antenna efficiency of 61.6% for Ag, 50.3% for Al, 34.3% for Au, 29.5% for Cu and 9.4% for Cr. In a later paper, [95], they looked at the diode efficiency and more specifically, investigated the power loss due to impedance matching whilst assuming efficient signal rectification. This analysis was only done on the highest performing materials – Ag, Al and Au. Their work was a continuation of a previous work by Sarehraz *et al.*[96], however their analysis covered 75% of the power radiated by the sun (up to 1000 THz) whereas Sarehraz *et al.* only based their analysis on 15% of the spectrum. Although Al gave the best matching efficiency of 97%, the best overall conversion efficiency of the rectenna considering both constituent parts is with Ag which gives 54% whereas Al would have 46% and Au just 28%[95](Figure 8). So far the analysed publications all share a united opinion that a suitable rectifier is needed in order for the above mentioned efficiency estimates to be achieved or improved. For impedence matching, Ma & Vandenbosch[95] suggest that the impedance of the rectifier must be of the order of 100Ω both for the real and imaginary part and favour aluminium as being a suitable material for this application.

A governing factor in rectifier performance is its RC time constant which defines the cut-off frequency, f_c. This is the upper limit at which the rectification process is efficient. Above the cut-off frequency, rectification is still possible however the strength of the rectified signal at the output of the MIM will drop by a factor of approximately $(f_c)^2$, thus the higher the frequency above f_c the lower the rectification efficiency[96]. The cut-off frequency is estimated by the following expression proposed by Sanchez *et al.*[104],

$$f_c = \frac{1}{2\pi RC} \tag{2}$$

where R is the series resistance and C is the capacitance. The cut-off frequency of the whole rectenna device can be evaluated by the following relation, which considers the antenna impedance (R_A) and diode resistance (R_D)[92],

$$f_c = \frac{R_A + R_D}{2\pi R_A R_D C_D} \tag{3}$$

To achieve a high cut-off frequency, the RC constant must be small, which can be obtained with a small junction area[104], and appropriate material selection and fabrication. MIM rectifiers, although the most promising rectifier technology, have not been demonstrated at high THz frequencies with the best MIM diodes operating only up to 150 THz ($\lambda = 2$ μm)[5] (more detail in Section V). Improving MIM fabrication technology and material system design would potentially yield the solution to the rectenna efficiency.

So far, the reviewed efficiencies were all based on classical approaches, not concerning thermodynamic limits to solar energy conversion. In terms of equilibrium, an antenna receiving power from any source and transferring it to the load must transmit the same amount of power back to the source[3]. If the power is extracted from the load, the reduction in its temperature will introduce a different approach to analysing the energy balance between the incoming, extracted and reradiated powers. In this case, Corkish *et al.*[3] state that the system would have the same conversion efficiency limit as the one expected for a solar thermal collector. Assuming the Sun to be a 6000 K black body and the surroundings to be at 300 K, this limit is 85.4% for mono-spectral energy conversion and 86.8% for multi-spectral. Further analysis suggests that the Landsberg[102] limit of 93.3% can be reached due to certain benefits that the rectifying diodes have in rectennas (see [3] for more details), however this is hard to justify due to the generation of thermal noise by the rectifiers. Diodes cannot rectify their own thermal noise as it violates the second law of thermodynamics (known as Brillouin's paradox[106]).

Lerner *et al.*[93] developed a thermodynamic expression

for the open circuit voltage and conversion efficiency of a rectenna with the rectification process based on the geometric property of the antenna tip which provides a connection to the circuit and creates a tunnel junction. They envisioned the possibility of using advanced nanofabrication techniques and a selective atomic layer deposition process to manufacture nanometer junctions which would potentially allow for rectification of blue light frequencies[93]. In modelling the circuit of the rectenna, it can either be considered as a constant voltage source with an emf independent of the load resistance or as a constant current source whose emf depends on the load resistance. As suggested by Lerner *et al.*[93], the rectenna device is more appropriately modelled as a constant current source in which case efficiencies are predicted in excess of 70% for a monochromatic light source and an optimised device. Should the device be tuned for multiphoton absorption, the efficiencies are expected to be further enhanced[93]. The question remains as to whether a suitable technology is yet available to make such structures.

Another analysis on the broadband blackbody radiation response of optical rectennas was done by Joshi & Moddel[90] by using the theory of photon assisted tunnelling. By assuming negligible reverse current of the diode, zero dark current turn-on voltage and perfect matching to the antenna, the device power conversion efficiency at monochromatic light can approach 100%, whereas for multispectral light the value is approaching a maximum at 44% at an operating voltage of 1.1 V. It should be noted that these authors also suggest that spectral splitting is likely to improve the efficiency. This can be achieved by setting the operating voltage for each rectenna to rectify the desired spectral range, thus implying that the operating voltage for the diodes in rectennas has a similar role to the band-gap as a limiting factor in photovoltaic devices[90].

With its proven success to receive and rectify monochromatic microwave beams with high efficiency, there is a growing interest in establishing the true potential of the rectennas' ability to convert solar radiation. As seen in Figure 8, there are many theoretical approaches predicting this, however there is no one universal theory that clearly states whether the solar rectenna can compete with, let alone surpass common photovoltaic cells. There is however, the common opinion between scientists worldwide that the technology is not mature enough to fabricate a device of comparable efficiency. Advances in antenna engineering suggest that it is now possible to manufacture nano sized antennas which can absorb solar radiation from the infrared to the visible frequencies (Section IV). Work on the rectifier component has seen great progress over the years with many attempts showing great potential for use in rectenna devices (Section V).

IV. ANTENNA

Antennas are used either as transmitters, to create electromagnetic (EM) waves that have a well-defined radiation pattern, or as receivers of EM waves from a remote source[86]. The purpose is usually to send or extract information that is encoded into the EM wave. Antennas are now essential in many aspects of modern day living, the mobile phone being the most common example. They normally operate at a resonance. Typically this is a half wavelength dipole antenna. The resonant wavelength of the antenna will be at the wavelength of the incoming propagating signal and energy is then transferred between them. For this reason, the antennas are usually fabricated in the dimensions of the wavelength of the incident wave. A higher frequency thus requires a smaller antenna. However, an antenna designed to be resonant at a very specific frequency by only considering a linear dimension does not guarantee efficient energy absorption. Other factors such as impedance mismatch in the circuits, bandwidth, direction and polarisation of incidence radiation needs to be considered. The aim of this section is to introduce some different designs of antennas for microwave and optical frequencies.

Most antennas at microwave frequencies are designed to work within a specific frequency band reserved for industrial, scientific and medical (ISM) frequency use. For example, one of the most popular frequency bands is 2.4 – 2.5 GHz, which devices such as wireless networks, near-field communications, and Bluetooth use. Therefore these antennas only require millimeter dimension. A dipole antenna is one of the simplest designs. It consists of two identical metal elements in mirror symmetry. The most common form is two metal rods aligned on the same axis. The length of the metal rods is half the size of the resonance wavelength. Higher-order resonance modes also occur in the same antenna but at reduced efficiency due to attenuation of the induced current in the antenna. Nevertheless, the accepted frequency bandwidth is very narrow. A slight variation of the incoming frequency gives rise to parasitic capacitance and inductance resulting in inefficient energy absorption. Therefore, a precise fabrication of the antenna is necessary. This increases the difficulty in manufacturing good devices, especially when the frequency of interest is higher.

One method to circumvent the narrowband absorption problem, which dipole antennas encounter is to extend the arms width above the standard limit (i.e. 5% longer than the wavelength)[108], allowing a bigger window of frequency acceptance. This type of antenna is referred to as the bow-tie antenna (Figure 9). The length of each arm is half the size of the radiation of interest. However, the correct length of the antenna alone does not guarantee high efficiency absorption since the incoming wave has various angles of incidence and polarisation[109]. To overcome this problem, a variety of designs have been investigated, such as spiral[108,110](Figure 9), log-periodic[108,110], microstrip slot[111], and retrodirective rectenna array[112].

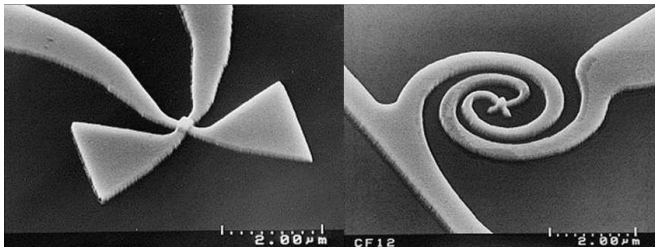

FIG. 9. SEM images of (left) bow-tie antenna with integrated Ni-NiO-Ni diode and (right) spiral antenna. (Reprinted from [108], ©1998 with permission from Elsevier)

As previously discussed in Section III, at microwave frequencies, antennas can be designed to be very efficient. Achieving such high efficiencies at the higher frequencies is more challenging with scaling down antennas. The solar spectrum emits energy across a broad band, which implies that the antenna must absorb a large bandwidth of solar frequency, otherwise most energy will be lost. For an antenna to operate efficiently at optical frequencies it is necessary to adjust both the inductance (L) and capacitance (C) to bring the device resonance to the required short wavelength. This would require both L and C to be very small[113]. To achieve this, the dimensions of the antenna must be reduced to the scale of the optical wavelength, i.e. nanometer. Scaling down the conventional antenna design to nanometer is of interest, as it can be determined whether classical antenna theory remains valid. For example, electron beam lithography has been used to fabricate a dipole antenna with a 3.5 nm NiO layer, sputtered between the two Ni antenna arms[109]. The dipole antenna had a total length of 6.7 μm and was designed for 10 μm wavelength (30 THz). Another example is of a gold dipole antenna fabricated by using focus ion beam (FIB) milling[114].

These attempts have proven it impossible to scale classical antenna designs to the optical frequency regime. At radio frequency the metals used for the construction of antennas can be considered almost lossless. This allows a large variation of radio frequency antennas to be designed by considering the metal to be a perfect conductor. However, the losses become increasingly significant as the frequency rises. This is due to the finite effective mass of electrons causing the electrons to react with increasing phase lag to an oscillating EM field[115]. As soon as the phase lag approaches 90^o the amplitude of the charge oscillation goes through a maximum and is only limited by the Ohmic and radiation damping of the system[116]. In metallic nanoparticles this resonance corresponds to the localized surface plasmon resonance (LSPR). For gold, silver, aluminium and copper the LSPR is in, or very close to the visible light part of the spectrum. The LSPR can be exploited to overcome some of the drawbacks of antennas at optical frequency, in particular the high Ohmic losses compared to microwave frequencies.

A. Properties of Metals at Optical Frequencies

The optical response of metals can be described by a complex frequency-dependent dielectric function,

$$\epsilon(\omega) = \epsilon_1(\omega) + i\epsilon_2(\omega) \tag{4}$$

The electric field is related to the induced polarization density as[117],

$$P(\omega) = \epsilon_0[\epsilon(\omega) - 1]E(\omega) \tag{5}$$

For optical antennas, Ohmic losses in the metal should be minimized. The Ohmic absorption is proportional to the conductivity of the material, $\sigma(\omega)$, which is related to the dielectric function by.

$$\epsilon_2 = \sigma(\omega)/\epsilon_0\omega \tag{6}$$

Ohmic losses take place in close proximity to the surface of the metal, within the so-called penetration depth[118], which for metals at visible wavelengths are of the order of several nanometers (approximately 13 nm and 31 nm for aluminium and gold at 620 nm wavelength, respectively)[119]. The dielectric properties of a metal can cause a LSPR in the visible spectrum, which is connected to large local fields, and enhanced scattering and absorption.

At optical frequencies the metal conduction electrons may be treated as an ideal electron gas. The collective behaviour of this free electron gas can be expressed using the Drude-Sommerfield model[118],

$$\epsilon(\omega) = \epsilon_\infty - \frac{\omega_p^2}{\omega^2 + i\gamma\omega} \tag{7}$$

where ω_p is the volume plasma frequency and γ is a damping constant[118]. (For Au at optical frequencies $\omega_p = 13.8 \times 10^{15} s^{-1}$ and $\gamma = 1.07 \times 10^{14} s^{-1}$)[116]

The DrudeSommerfield model does not account for interband transitions due to photons with high-enough energy promoting electrons from lower lying valence bands to higher energy conduction bands[120]. This can be described using a Lorentz model of the dielectric function, which is described by a collection of damped harmonic oscillators with well-defined resonance frequencies, ω_0,

$$\epsilon(\omega) = \epsilon_\infty + \sum_{i=1}^{n} \frac{\alpha_i \omega_p^2}{\omega_{0i}^2 - \omega^2 - i\gamma_i\omega} \tag{8}$$

where ω_p depends on the density of bound electrons involved in the absorption process and γ is a damping constant for the bound electrons. This Lorentz model shows strong deviation from the free electron gas model near ω_0, leading to a maximum in the imaginary part of $\epsilon(\omega)$ and strongly increased damping.

Au, Ag, Al and Cu are used as materials for metallic optical antennas. The dielectric constants of Au and

Cu are very similar, with a Drude-like response for wavelength above 600 nm and an onset of interband transitions occurring around 530–550 nm, making them excellent for antennas in the red and near-IR spectral region. For Ag the first interband transition is for a short wavelength, less than 400 nm, making it superior to Au for wavelengths around 500 nm. Al has a larger negative real part of dielectric function, and so among the four metals is the one that best approximates an ideal metal, especially in the 400–600 nm spectral region[116]. Unfortunately there is an interband absorption peak located at 800 nm wavelength, making Al unsuitable for use in the near-IR region, although, as previously discussed, it makes a good candidate for solar rectenna due to its high predicted matching efficiency[95]. As well as the spectral properties of the metals, the chemical stability also needs to be carefully considered. Ag and Cu are known to rapidly oxidise under ambient conditions and Al forms thin passivation layers of Al_2O_3. Due to its dielectric functions in the red and near IR parts of the spectrum and its excellent chemical stability, Au is the material most often used as a nanoantenna.

B. Scattering and Absorption Properties of Optical Antennas

Radio frequency and microwave antennas will normally always be considered as circuit elements connected to a feeding circuit. At optical frequencies the circuit element models are not valid and so optical antennas often appear as isolated or electromagnetically coupled structures, whose resonant properties have to be considered. Here, the LSPR of single and coupled metal nanoparticles, specifically the absorption and scattering properties, are considered. Consider a monochromatic plane wave of irradiance I_i incident on a sphere of radius a. If the irradiance is multiplied by a cross-section then the power absorbed or scattered by the particle can be calculated,

$$P_{abs} = I_i.C_{abs} \tag{9}$$

$$P_{sca} = I_i.C_{sca} \tag{10}$$

where C_{abs} and C_{sca} are the absorption and scattering cross-sections, respectively. The cross-sectional areas can be normalized to the cross-section of the sphere to give the scattering and absorption efficiencies:

$$Q_{abs} = \frac{C_{abs}}{\pi a^2} \tag{11}$$

$$Q_{sca} = \frac{C_{sca}}{\pi a^2} \tag{12}$$

where a is the radius of the sphere. Equations 11 and 12 define the two loss mechanisms, Q_{sca} which is radiative and Q_{abs}, which is non-radiative, the extinction efficiency being the sum of both these terms.

Smaller sized particles restrict electron motion and cause electrons to collide with the inner surface of the particle, resulting in loss of energy and increased absorption. Absorption is therefore the greater loss mechanism in smaller particles. The radiation damping rate is proportional to the number of electrons. This means that the larger the particle, the more significant this factor and the broader the plasmon resonance becomes. A larger particle volume results in a decrease in the amount of incident radiation absorbed but an increase in the scattering. Scattering and absorption due to small spheres can be described using the well-known Mie theory[121], where

$$C_{abs} = \frac{2\pi n_1}{\lambda_o} Im(\alpha) \tag{13}$$

$$C_{sca} = \frac{\left(\frac{2\pi n_1}{\lambda_o}\right)^4 |\alpha|^2}{6\pi} \tag{14}$$

The polarization, α, is given by:

$$\alpha = r^3 \frac{(\epsilon_2 - \epsilon_1)}{(\epsilon_2 + 2\epsilon_1)} \tag{15}$$

where r is the radius of the sphere, ϵ_2 the complex permittivity of the sphere and ϵ_1 the complex permittivity of the surrounding medium. Examples of Mie theory calculations of extinction, scattering and absorption for gold and silver spheres of various radii are shown in Figure 10.

Interesting observations that arise from the Mie theory calculations are that a number of peaks can be observed for larger diameter spheres, the longest wavelength peak being a dipole mode. As expected the LSPR broadens as the particle size increases but absorption and scattering maximums are not always at the same wavelength. The absorption and scattering properties can be tuned using different shaped particles. In this case more complex calculation methods are required, for example the Discrete Dipole Approximation (DDA)[123], Finite Element Method (FEM)[124] or Finite Different Time Domain (FDTD)[125]. FDTD is a popular method since it is possible to consider the scattering and absorption from any shaped particle in any dielectric environment and since the calculation is in the time domain a single calculation will yield the spectral response over a wide bandwidth, using a Fourier Transform[126].

If a surface that fully encloses the scattering object is now defined, then by considering the total field flowing through it, one can calculate the normalized absorption cross-section Q_{abs} from[127],

$$Q_{abs} = \frac{\int\limits_s Re(0.5\mathbf{E} \times \mathbf{H}^*) d\mathbf{a}}{|S_i|} \tag{16}$$

and by considering the scattered field the normalised scattering cross-section, Q_{sca}, is found from[127],

$$Q_{sca} = \frac{-\int\limits_s Re(0.5\mathbf{E}_s \times \mathbf{H}_s^*) d\mathbf{a}}{|S_i|} \tag{17}$$

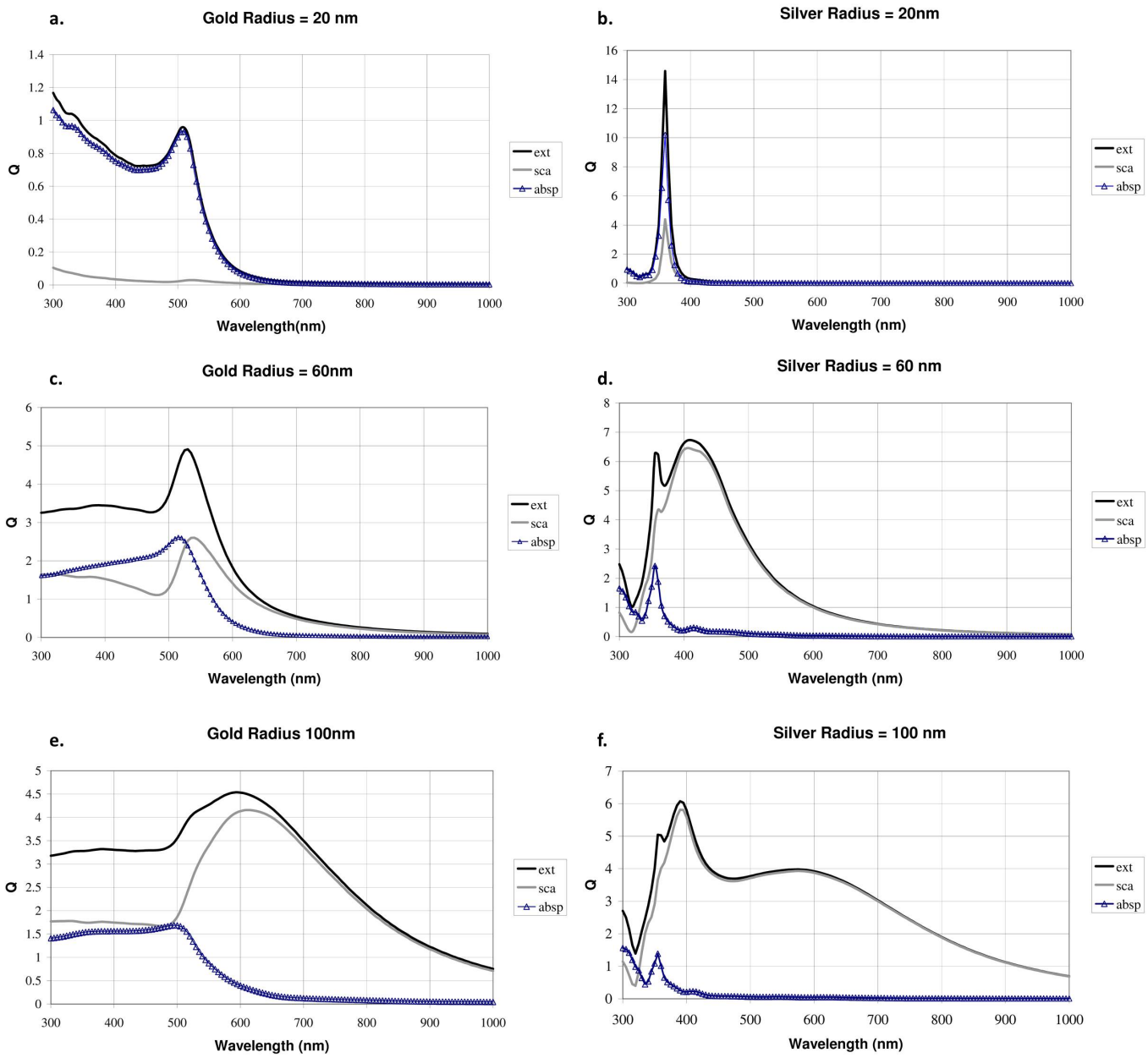

FIG. 10. Mie calculations of extinction, scattering and absorption efficiencies for silver and gold nanospheres (a) Au sphere of 20 nm radius (b) Ag sphere of 20 nm radius (c) Au sphere of 60 nm radius (d) Ag sphere of 60 nm radius (e) Au sphere of 100 nm radius (f) Ag sphere of 100 nm radius.

where $|S_i|$ is the power incident on the cross-sectional area of the nanoantenna. Using computational electromagnetics to calculate the electric (**E**) and magnetic (**H**) fields at the enclosing surface, Equations 16 and 17 can be solved. Figure 11 shows the Q_{sca} calculated using FDTD for (a) two closely spaced and (b) two overlapping spheres, where the spheres are silver and of 50 nm diameter[122].

Furthermore, the surrounding dielectric environment and orientation of the nanoparticle can significantly modify the absorption and scattering properties[127]. As an example, consider the two configurations of 30 nm radius hemispherical Au nanoparticles, depicted in Figure 12. In both configurations the metal nanoparticle is at the interface between the air and silicon on the lower surface of the semiconductor. In the first case the nanoparticle extends into air whilst for the second case it extends back into the silicon. In each case the scattering properties of the metal nanoparticle are calculated when there is a normally incident field from within the silicon[127]. The results are shown in Figure 13. It can be seen that in case (b) the scattering efficiency is dominant whereas in

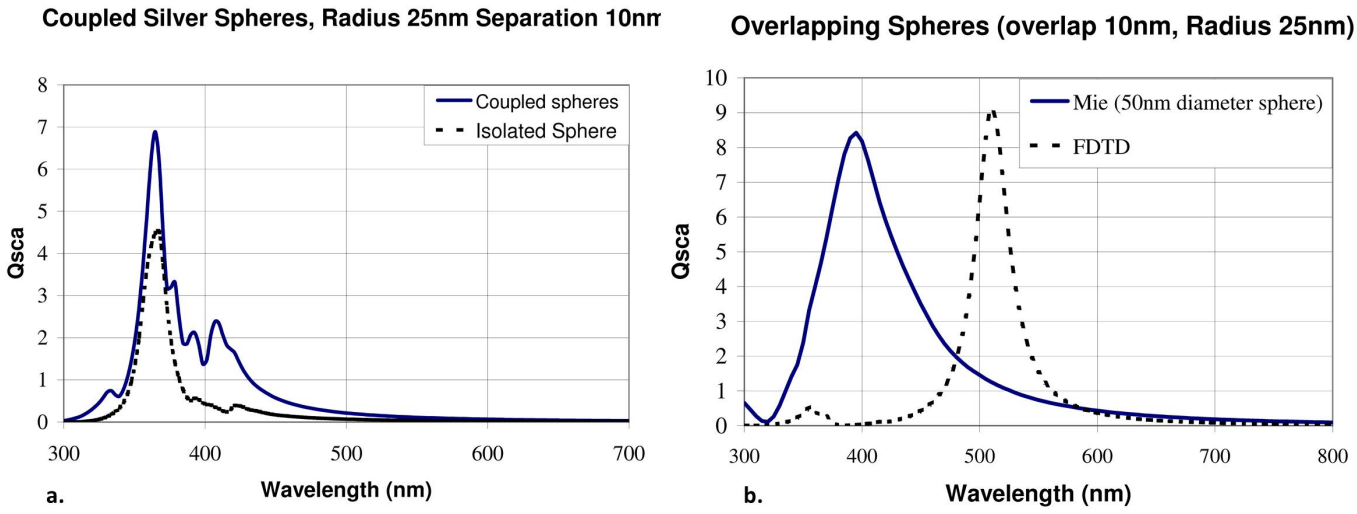

FIG. 11. FDTD calculations of Q_{sca} for (a) two closely coupled Ag spheres and (b) two overlapping Ag spheres.

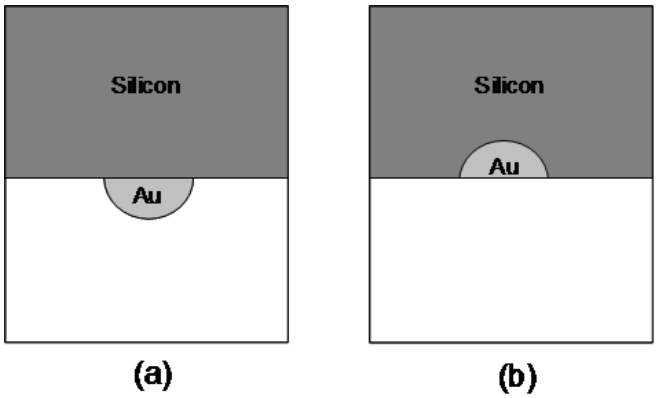

FIG. 12. Depiction of the two configurations considered: (a) nanoparticle extending into air and (b) nanoparticle extending into silicon. (Reprinted from [127])

case (a) it is the absorption that is the largest extinction component.

We have discussed the LSPR of isolated and coupled nanoparticles in terms of their absorption and scattering efficiencies. The subsequent coupling between the nanoantenna and rectifying diode will need to be optimised. Absorption is caused by the generation and recombination of electron hole pairs (EHP). EHP formation occurs when the plasmon quantum excites an electron from the Fermi level to an occupied state below the vacuum level. It has been shown that the concentrated electric fields around a nanoantenna, due to the LSPR, can cause hot electron carriers[128–130]. Because of their higher energy, hot electrons will extend further away from the nanoparticle than an equilibrium electron distribution, which is above the vacuum level. If a nearby electron acceptor is present, hot electrons can transfer into its electronic states. This hot electron carrier generation and injection into an acceptor due to the high fields is a topic of much current research in photovoltaic devices[128–131], in particular optimizing the efficiency of this process.

For an optical rectenna it is desirable to enhance and localize (i.e. concentrate) the propagating fields to enable coupling into external circuit elements. The concentrated fields tend to be higher for particles with a higher scattering efficiency and between the small gaps of coupled nanoparticles. This is opposite to hot electron devices which seek to maximise the absorption of photons in the nanoparticle. Figure 14 shows the values of electric field, normalized to the incident field, around Ag nanoparticles, formed by colloidal lithography using 500 nm polystyrene spheres as a template on a glass substrate[132] (this makes the triangular sides around 120 nm in this case). The scale is logarithmic and it can be seen that there are up to two orders of magnitude electric field enhancement in the gaps between the particles.

C. Alternative structures

Progressing toward a more efficient energy harvesting technology, some creative and novel designs have been explored. Other than the hot electron assisted mechanism, the idea of optomechanical nanoantenna has also been explored. In this design, the plasmonic bowtie antenna when coupling with the incident radiation, generates a strong electric field in the gap regions of the antenna arms and bend the arms toward each other[133] (Figure 15). As the arm bends toward each other, a redshift in transmission and reflection power occurs, allowing the device to have a larger operation bandwidth. With the inclusion of surface plasmon polariton property, it is calculated that

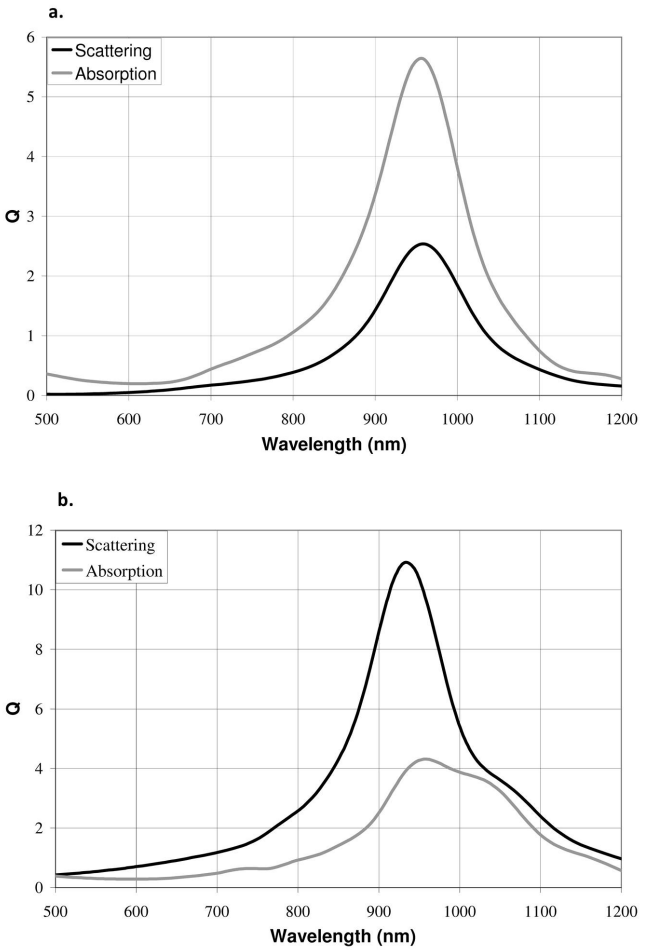

FIG. 13. Q_{sca} and Q_{abs} FDTD calculations for (a) nanoparticle extending into air (b) nanoparticle extending into silicon. (Reprinted from [127])

the antenna length can be greatly reduced while provide a mechanical bandwidth of 4.4 GHz.

The propagation wavelength for the waveguide in the material is very small. This suggests that if the antenna is very close to the diode, it can reduce losses[116]. In Travelling Wave (TW) MIM rectenna design, a MIM junction is created between the top and bottom bow-tie antenna[134,178] (Figure 16a). This allows the diode to rectify the absorbed radiation at this MIM junction. This design is similar to a bow-tie antenna, except that the left bow and the right bow are separated by a very thin layer of insulator, forming an extended MIM tunnelling diode. The absorbed radiation excites surface plasmon and propagates along this axis (Figure 16c). The current induced between the top metal and bottom metal cause tunneling of electrons to occur. A final rectified current is produced due to the asymmetry in the IV characteristic at the bias point (further information in Section V). One of the advantage in this design is that it is not limited by the RC bandwidth.

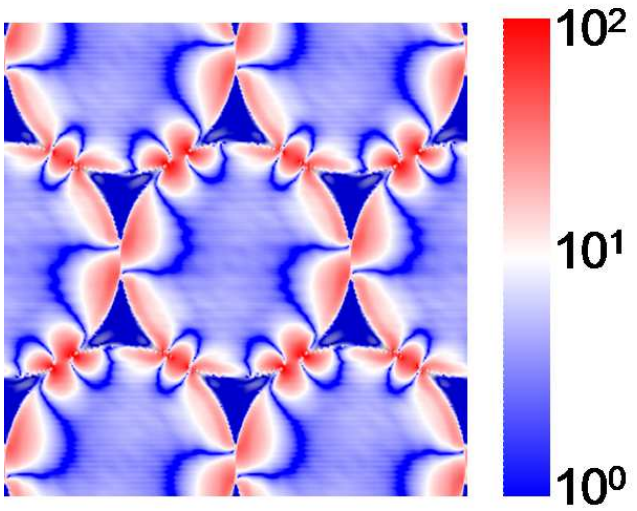

FIG. 14. Concentrated electric field calculations around triangular like nanoparticles of height 100 nm formed by colloidal lithography with 500 nm polystyrene spheres. The excitation wavelength is 780 nm and calculations were carried out using FDTD. (Reprinted from [132])

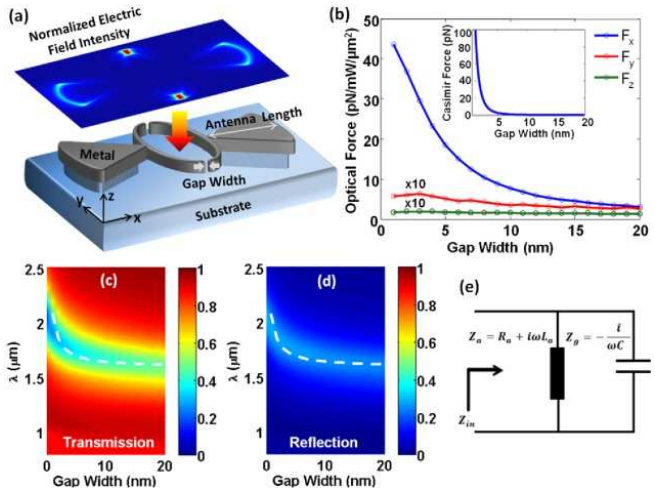

FIG. 15. (a) 3D schematic diagram of the optomechanical nanoantenna investigated. A near-field intensity map of the device is shown on top. (b) Simulated optical force and Casimir force (inset) on the suspended beam as a function of gap width. (c) Simulated (a) transmission and (b) reflection power spectrums versus gap width. (e) Nanocircuit lump elements model. (Reprinted from [133], ©2012 with permission from Optical Society of America)

Another example of integrated rectenna design is the slot-antenna-based frequency selective surface (FSS) with integrated MOM diode[135]. The materials consists of Al, Al_2O_3 and Pt on benzocyclobutene (BCB) standoff, designed for absorbing radiation at 28.3 THz (10.6 m). Numerical simulations showed that when the incoming radi-

ation is parallel to the diode junction, the electric field is concentrated 2000 times across the diode with a decrease of reflectance around the desired wavelength, which can be used as an indication of absorption. A working device was fabricated with e-beam lithography (Figure 17) and similar optical response in reflectance was measured.

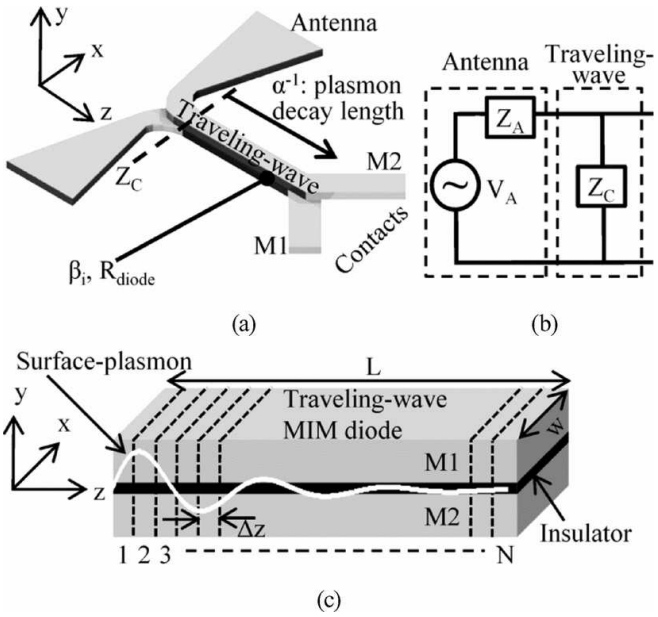

FIG. 16. (a) Isometric view of the antenna-coupled TW device. Instead of crossing over as in a typical bow tie antenna in figure 1, the antenna arms converge into a parallel-plate waveguide with a 2 nm thin insulator between the metal M1 and M2. (b) Circuit representing the TW MIM device. (c) 3D view of the TW MIM diode. (Reprinted from [134], ©2010 IEEE. Reprinted, with permission, from IEEE)

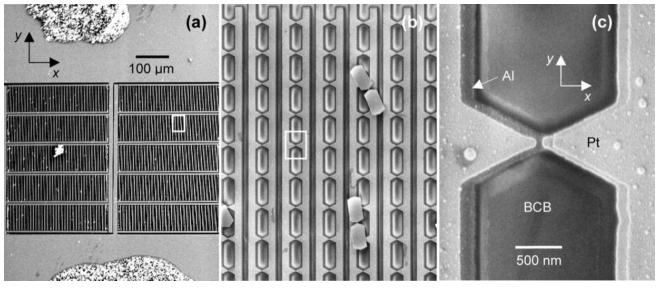

FIG. 17. SEM images of fabricated MOM diode-coupled frequency selective surface device.(a) Image of entire array consisting of 16,800 diode-coupled elements is less than 0.325 mm^2, (b) closer image of the array showing several pieces of metal coated photoresist that became attached to the sample after lift-off, and (c) diode of working device (Reprinted from [135], ©2012 with permission from John Wiley & Sons)

The recent development of graphene technology also prompts the incorporation of graphene into nanoantenna design. Ag is known to suffer from sulfidation under ambient environment, resulting in degradation of the plasmonic property[136]. Graphene was shown to be able to passivate the Ag and prevent sulfidation without affecting the plasmonic property of the Ag nanoantenna[136]. Additionally, radical changes in the plasmon energy and strength were also observed by electrically doping patterned graphene arrays with a gate voltage[137]. The development of new materials such as graphene open up new possibility to design a better optical antenna.

D. Impedance matching

As previously introduced, the efficient energy transfer from antenna to diode requires an impedance match between both components. A theoretical analysis on an optical dipole nanoantenna showed that the nanoantenna appears to be an effective RLC parallel circuit[138] (Figure 18). The two parallel resistances are due to radiative decay and non-radiative decay rates. The electrical properties of a nanoantenna are intrinsically influenced by the material used, dimensions and geometric shape[95]. Dipole antennas of length between 100 and 350 nm for Ag, Al, and Au on glass have been studied through simulation[95], as previously introduced. At a single frequency and assuming no loss of power in impedance mismatch, a silver dipole shows the highest efficiency. However, the upper bound efficiency drops when the solar irradiance spectrum is taken into account. Further investigation by inclusion of losses due to impedance mismatching showed even further decrease in conversion efficiency (see Section III).

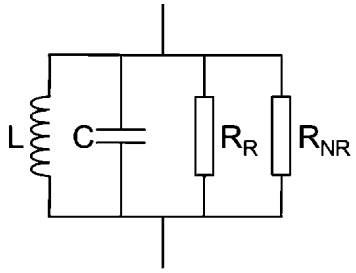

FIG. 18. Equivalent circuit of a metallic nanoantenna. (Reprinted (Fig. 1) with permission from [138], ©2010 by the American Physical Society)

Liu *et al.*[139] considered the nanoantenna as lumped elements in three-dimensional nanocircuits and demonstrated that impedance of the antenna can be tuned by loading a dielectric or a metal in the gap between a dimer. A SiO_2 load was used as the dielectric material. In the circuits it represents an additional capacitor for the dimer system. When gold is loaded between the gold dimers, the nanocircuit becomes an LC circuit. The gold load acts as an optical frequency inductor and the two gaps between the gold load and gold antenna behave as two

capacitors. To develop the idea further, a combined gold-SiO_2 load was also fabricated. The circuit now corresponds to an LC parallel circuit in series. In all cases, the impedance was tunable by changing the dimensions and the materials of the load, with the gold-SiO_2 showing largest tunability.

E. Nanofabrication

For microwave energy extraction device dimensions are in the micrometre range, achievable using well established photolithography techniques. At infrared and optical frequencies nanometre dimensions are required. Unfortunately the resolution of photolithography is limited by the diffraction wavelength of light and so different fabrication techniques are required. Fabrication techniques that have been demonstrated to be effective at obtaining nanometre resolution are electron beam lithography, nano-imprint lithography and colloidal lithography.

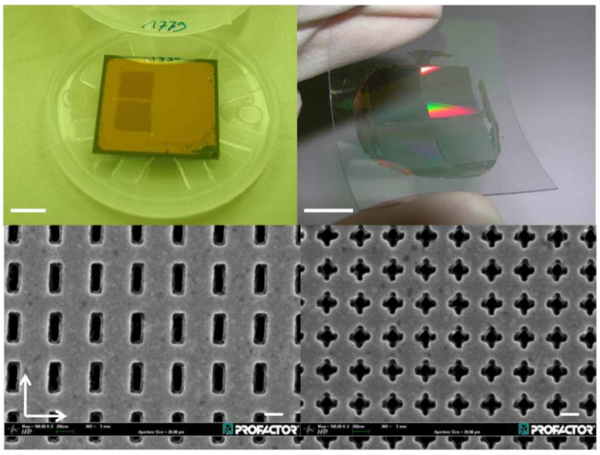

FIG. 19. (a) Photograph of a single layer negative index material (NIM) sample and (b) transfer printed gold grating on a flexible foil. Top-view SEM images of (c) the single functional fishnet layer and (d) the Swiss-cross NIMs fabricated using nanoimprint technology. (Reprinted from [141], ©IOP Publishing. Reproduced by permission of IOP Publishing. All rights reserved.)

Electron beam lithography is similar to photolithography but uses an electron beam instead of light. The wavelength of the electron beam is much smaller than light, allowing sub-10 nm resolution in patterning and creation of well-defined structure such as the nanogap in antenna-coupled diode in Figure 18. However it is costly and cannot be effectively up scaled for a large area. Nanoimprint lithography uses a pre-made mask to create a pattern on a substrate[140]. This mask can be used numerous times and allows large area fabrication. Compared to e-beam lithography, nanoimprint lithography enables large area patterning but each mask is specially designed for a specific nanostructure. Figure 19 illustrates an example of a large area of nanostructures fabricated using nanoimprint technology. A new mask will have to be made if the nanostructure required is different and the cost of the mask is high. Nanosphere lithography is a more versatile lithography technique, based on self-assembly of nanospheres. It allows fabrication of large area devices at relatively low cost. Although it does not allow huge degree of manipulation in the shape of nanostructure compared to e-beam lithography, a variety of shapes in large ordered array has been fabricated successfully (Figure 20), such as nanotriangles[132,142,143], crescent moon[144], nanorings[144], nanocones[145], and three-dimensional structures[146].

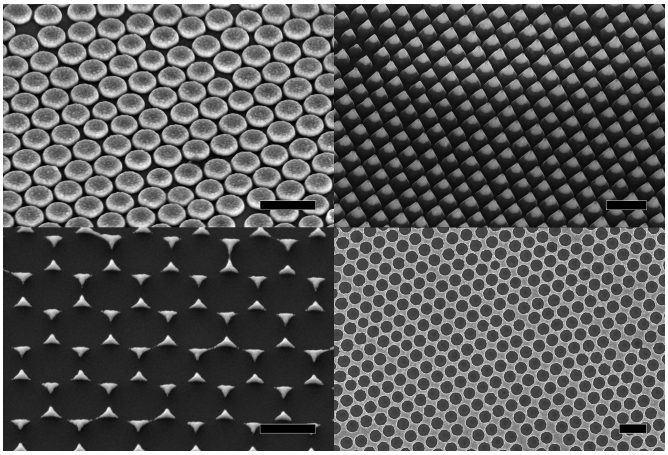

FIG. 20. SEM images of variety of gold nanostructures based on nanosphere lithography technique (a) nanotriangles, (b) slanted nanocones, (c) nanodisc array, (d) nanoholes. (*Note: 500 nm scale bar in all images*)

Advances in nanofabrication techniques are crucial in developing a better nanoantenna. Further details on nanoantenna theory and application can be found in some recent reviews [116,147].

V. RECTIFIER

A rectifier, a non-linear device such as a diode, is an integral part of the rectenna device, used to convert an AC input voltage from the antenna into a usable DC voltage. As discussed in Section III, the rectenna is fundamentally limited by the cut-off frequency (f_c) of the diode, above which the rectification process will become increasingly inefficient. f_c changes with the type of rectifier used due to the physical nature of the diode's operation dictated by its potential of achieving a low RC time constant. Schottky barrier diodes have been demonstrated and are theoretically limited to frequencies of a few THz, whereas MIM/MIIM diodes are shown to be operational at 150 THz[5,148] and are predicted to potentially exceed this and reach the visible regime. Three

parameters have been defined[5,149,185,188], which, if satisfied, result in characterising a device as a diode rectifier:

$$f_{asym}(V) = \left|\frac{I_F(V)}{I_R(V)}\right|; \textbf{ Asymmetry>1} \quad (18)$$

$$f_{NL}(V) = \frac{dI}{dV}(V)/\frac{I(V)}{V}; \textbf{ Nonlinearity>3} \quad (19)$$

$$f_{RES}(V) = \frac{d^2I}{d^2V}(V)/\frac{dI}{dV}(V); \textbf{Responsivity>7}V^{-1} \quad (20)$$

These three Figures of Merit (FOM) are commonly used to characterise the performance of MIM devices however can be equally applied to other diodes. The asymmetry FOM, defined in Equation 18, is the absolute ratio of forward to reverse current at a bias, a value of 1 indicating full symmetry and hence no rectification. The non-linearity FOM, as defined by Equation 19, is the ratio of the differential conductance to the conductance and is a measure of the deviation from a linear resistor. Finally the responsivity FOM defined in Equation 20 is the ratio of the second derivative of the I-V curve and the differential conductance and is a measure of the rectified signal as a function of input power. It is essential that the diode in a rectenna exceeds all of these criteria as the higher the FOMs for the diode, the more efficient the rectification process.

note: There was some disagreement argued by Tucker & Feldman[150] that responsivity should be described using second-difference equations and not derivatives in order to understand high frequency rectification. However, proving which is the correct approach is beyond the scope of this review as the authors aimed to only bring this theory to the attention of readers. The authors support the use of derivatives of the I-V curve in analysis of these FOMs.

The fundamental operational theory and recent experimental and theoretical developments of Schottky barrier diodes (used in the microwave power transmission rectenna), metal-insulator-metal and metal-insulator-insulator-metal diodes (proposed for solar rectenna), and geometric diodes (proposed for solar rectenna) will be reviewed in this chapter.

A. Schottky Barrier Diodes

Schottky barriers have been used as rectifying devices since the first cat's whisker detectors in the first decade of the 20$^{\text{th}}$ century. The 'barrier' itself is formed within the semiconductor side of a metal-semiconductor (MS) interface, the result of its conduction and valence bands bending to align the Fermi levels of each material. Figure 21 shows an ideal MS interface with the Schottky barrier, $\Phi^0_{B,n}$, marked.

In Figure 21, note that within the semiconductor, a depletion region, or space charge region, forms. This is a region of the semiconductor that is devoid of mobile charge carriers when the device is unbiased, and as such it can be considered as the dielectric between two plates of a capacitor, so having a direct influence on the *RC* time constant. Minimising the capacitance of a Schottky diode can be achieved through scaling (though this has the equal and opposite effect on the diode's resistance), or by increasing the depletion region width by using low semiconductor doping.

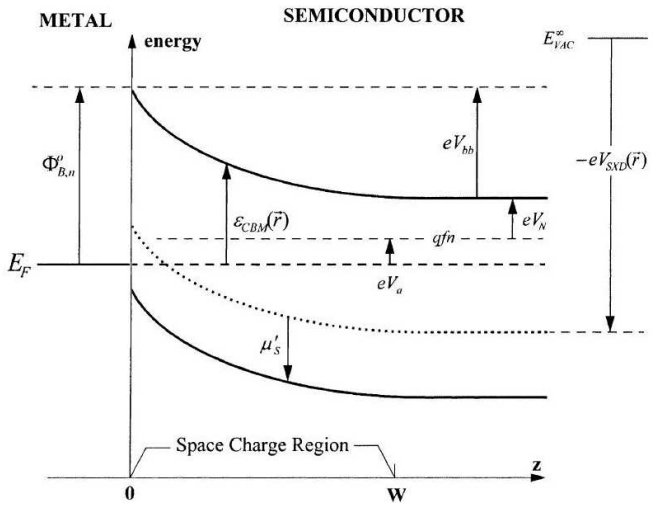

FIG. 21. Band diagram of an ideal Schottky Barrier, $\Phi^0_{B,n}$, contact formed with n-type semiconductor with an applied bias, V_a. (Reprinted from [151], ©2001 with permission from Elsevier)

The full operation of a diode has been well studied and understood over the past century, and readers are referred to [152,154], for a basic understanding of Schottky diode formation, operation and physics under different biasing conditions. Both the forward and reverse current-voltage (I-V) relationship is in most cases, where semiconductor doping is low enough, dictated by thermionic emission, rather than recombination or tunneling. The equation for thermionic emission, where the carrier in the semiconductor must pass up and over the potential barrier into the metal, is classically expressed as[152–154],

$$I = AA^*T^2 \exp\left(-\beta\Phi^0_{B,n}\right)\left[\exp\left(\beta V_A/\eta\right) - 1\right] \quad (21)$$

where A represents the contact area, A^* is the Richardson constant, T is the temperature, V_A is the applied voltage, η the ideality factor and $\beta = q/k_bT$, with q the electron charge, and k_b the Boltzmann constant. However, built into Equation 21 is the presumption that a perfect, homogeneous interface exists, with just one uniform value of $\Phi^0_{B,n}$. Instead, a more realistic scenario should be considered where multiple current paths exist flowing over barriers of different barrier height, so modifying this equation that has been the standard for a century or more. The papers by Tung[151,160] and others[155–159] pro-

vide a much more rigorous understanding of the Schottky diode and its operation under this inhomogeneous regime. Given the small potential size of a Schottky rectenna, the impact of such interface inhomogeneity will be the source of significant device-device variation, and an unpredictability of cut-off frequencies.

Typical Schottky diodes can have good diode characteristics in terms of the FOMs mentioned in Equations 18, 19 and 20, as demonstrated by the simple example of a Cr/Si Schottky diode made in-house for [155]. The characteristics of this diode are close to ideal (i.e. $\eta < 1.1$), and its FOM values are strong, with a diode asymmetry of 1669, nonlinearity of approximately 0.2 and responsivity above 10 V^{-1}. However, high series resistance and depletion region capacitance result in a cut-off frequency of just 40 GHz. Therefore, this is a device which could be suitable in a microwave power transmission rectenna, but it would be entirely ineffectual at solar frequencies.

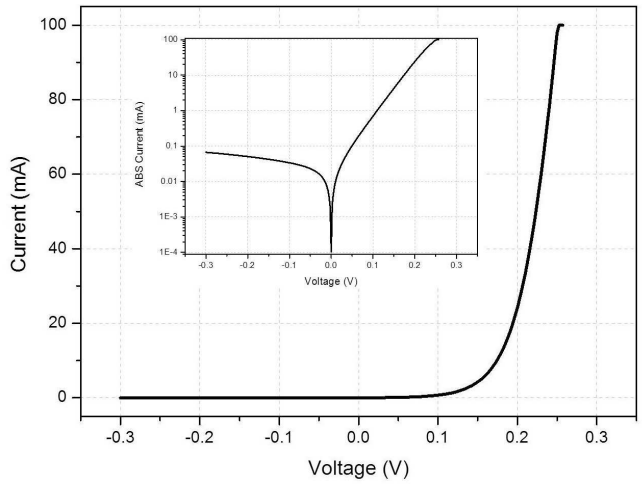

FIG. 22. Room temperature Current-Voltage characteristics of a simple Cr/n-Si lateral type Schottky barier diode of area $0.448mm^3$. SBH calculated at $0.51eV$, ideality factor of 1.06, diode asymmetry at ± 0.25 V of 1669, nonlinearity estimated at 0.2 and responsivity above 10 V^{-1}. The cut-off frequency has been estimated at 40 GHz.

With appropriate device design in terms of size and interface homogeneity, parasitic capacitances and resistances, arising from skin effects and residual native oxide scales at the MS interface, can be minimised to have a significant effect on the RC time constant. In silicon the cut-off frequency has been seen to reach 1 THz obtained by a layer of Ti-Pt-Au on a thin layer of n-Si[161], whereas 400 GHz has been seen from a simple Ti/n-Si structure[162]. On GaAs, frequencies of up to 5 THz have been achieved, as reviewed by Sizov & Rogalski[163]. This is obtained using highly doped GaAs substrate with an Ohmic contact on the back side upon which a thin (0.3-1 μm) epitaxial GaAs layer is grown, with Pt forming the Schottky contact. A honeycomb diode chip design enables thousands of diodes (diameters of 0.25-1 μm) on a single chip whilst minimising parasitic losses.

As rectifiers in the wireless power transmission rectenna application, SBDs are more than suitable candidates and as seen in Section III can operate efficiently. With cut-off frequencies in the low THz, Schottky diodes may potentially find usage as infrared rectennas although they will be limited to only function at a fraction of the far-infrared range, where the incident solar radiation is only up to 0.3 $W/m^2/nm$. The aim is to have solar rectennas operating at the high THz region where solar radiation is up to 1.7 $W/m^2/nm$, however Schottky diodes will not be able to efficiently rectify such high signals. Therefore, other approaches such as the MIM/MIIM and geometric diodes are being considered for the solar rectenna.

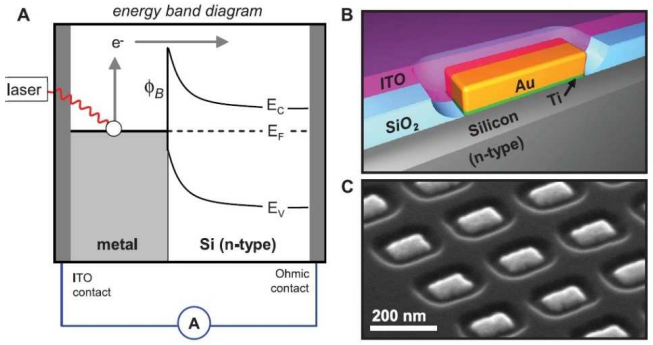

FIG. 23. Schottky diode as the active component for the solar rectenna device - concept inspired by Knight *et al.* **A** showing a band diagram of the proposed photocurrent mechanism, **B** showing the device layout and structure and **C** showing an SEM image of the array of devices. (Reprinted from [94], ©2011 with permission from The American Association for the Advancement of Science)

An alternative approach, which was first proposed by Knight *et al.*[94], is to use the diode as the active component in a hot-electron device. In this case, the array of nano-metre scale anodes double as the antennas, in which the incident light causes a localised surface plasmon resonance. Plasmon decay produces electron-hole pairs and as represented in Figure 23A, hot electrons are injected into the semiconductor after overcoming the Schottky barrier energy. The reported[94] quantum efficiency of this structure is estimated at being just 0.01%, though improvements are suggested to reach efficiencies of 2%. Recent publication suggests 30% efficiency of converting "hot electrons" in the metal to electrons in the semiconductor[164], though full conversion efficiency has not been estimated. In this case, although the device is based on an antenna coupled to a rectifier (similar components to a rectenna), the device is not proposed for light harvesting purposes due to its low efficiency and is instead demonstrated as a photodetector. Development of this concept into an energy harvesting applications

would require the conversion efficiency to be increased, though with the current state-of-art, this seems improbable.

B. Metal-Insulator-Metal Diodes

Current transport in Metal-Insulator-Metal (MIM) diodes is due to quantum electron tunnelling through an ultra-thin insulator layer (typically < 10 nm[165]). Rectification in these devices is due to variation in tunnelling rates caused by barrier asymmetry due to material selection, device geometry, thermal asymmetry, and photon induced deviation in electron flux distribution across occupied and unoccupied states[105], all contributing towards a nonlinear asymmetric I-V response. In the current review, we will qualitatively explain the nature of the I-V asymmetry taking into account the material selection for planar MIM devices under no illumination and thermal equilibrium. Under illumination, electron tunnelling is stimulated by different processes depending whether an antenna is or is not coupled to the diode. Further reading is suggested in [105,165,166,169]. Thermionic emission also contributes to tunnel diodes' I-V curves[165,188,189], especially to devices with thick insulating layers[188].

1. *Theory of operation*

Consider a Metal-Insulator-Metal junction as shown in Figure 24**a**. Metal 1 has a workfunction, Φ_1, lower than that of Metal 2, Φ_2, and χ is the electron affinity of the Insulator. The barrier height at the metal-insulator interface is determined from the difference between the workfunction and electron affinity respectively. At zero bias the band structure is shown in Figure 24**a**. By applying a negative bias (Reverse bias) to Metal 1, the band diagram shifts as in Figure 24**b**. In this case direct tunnelling is observed as a result of the *effective* tunnelling distance, d_t, being the same length as the insulator thickness. Further increasing the applied voltage, V_A, decreases the *effective* d_t, and it becomes shorter than the length of the insulator (as seen in Figure 24**c**). In this case, Fowler-Nordheim tunnelling[167] starts to occur. When a bias of the same magnitude is applied to Metal 2 (Forward biasing), the bands shift as in Figure 24**d**. Fowler-Nordheim tunnelling is observed in this case and occurs at a lower voltage, the higher the barrier asymmetry at both metal-insulator interfaces (arising from the difference in metal workfunctions). Further increasing the reverse bias (as in Figure 24**e**) will further decrease the *effective* tunnelling distance, d_t, which further increases the tunnelling probability in forward bias. Therefore, rectificaton occurs because for the same voltage magnitude in forward bias, d_t will be shorter and more current will be flowing than in the equivalent case in reverse bias. By changing the applied voltage, the tunnel distance changes linearly and the tunnel current is ex-

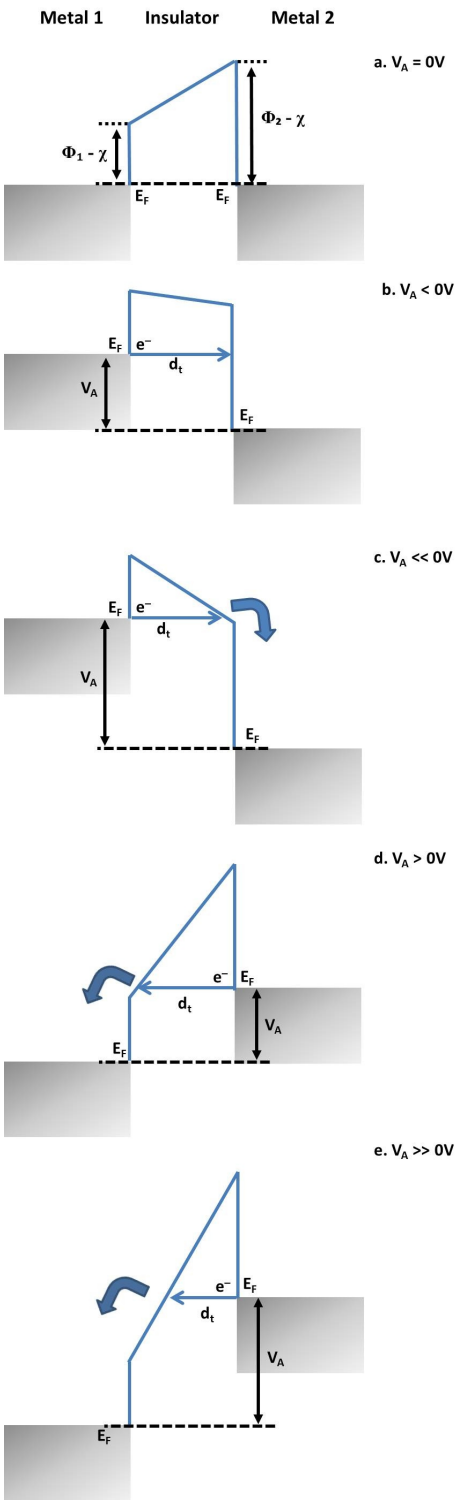

FIG. 24. Ideal band diagram for a MIM diode with dissimilar workfunctions for Metal 1, Φ_1, and Metal 2, Φ_2, χ the Insulator's electron affinity, E_F the Fermi level, d_t the tunnel distance and V_A the applied voltage. **a.** No bias situation ($V_A = 0V$); **b.** & **c.** Negative bias at Metal 1 (Reverse bias); **d.** & **e.** Positive bias at Metal 1 (Forward bias).

ponentially dependant on this distance[168]. This leads to an asymmetric, nonlinear current-voltage response and is the origin of the rectification mechanism in MIM diodes.

2. *Experimental devices*

Various metal-insulator systems have been used to investigate the rectification behaviour of MIM diodes. Research attempts are concentrated either towards modelling the behaviour and rectification properties or improving the figure of merits by optimising the material selection and fabrication techniques of these structures. As theory suggests, using similar metals in the MIM configuration should yield symmetric I-V curves, however asymmetric I-V curves have been reported for $Al/Al_2O_3/Al$[170] with structures and experiments conducted by the authors on $Ti/TiO_2/Ti$ and $Nb/Nb_2O_5/Nb$ devices having also displayed asymmetric I-V curves for these structures. The origin of this asymmetry is suspected to come from the formation of two different metal-insulator interfaces arising from dissimilar deposition processes, thus creating regions of enhanced leakage currents through the barrier. Other reported structures with optimised oxide layers such as Ni/NiO/Ni[108,174,176–178,180], $Al/Al_2O_3/Al$[171] and $Nb/Nb_2O_5/Nb$[149,185,187] show symmetry in the I-V curves, which is compliant with theoretical predictions. Although not displaying rectification properties, the Ni/NiO/Ni system is widely used for detection and mixing up to 30 THz infrared radiation due its significantly low resistance-area product of 1 $\Omega(\mu m^2)$[108,176–178], which indicates the potential for a low RC constant. However due to a lack of asymmetry, the second Ni electrode must be replaced with a metal of different workfunction in order to achieve rectification. This has been investigated by Hoofring *et al.*[175], Krishnan *et al.*[111,179] and Esfandiari *et al.*[181] with Au, Cr and Pt contacts respectively, all displaying good asymmetrical I-V response. Other structures of dissimilar materials which have shown promising results are $Al/Al_2O_3/Ni$[172], $Nb/Nb_2O_5/Pt$[149,184,185], $ZrCuAlNi/Al_2O_3/Al$[182] and also $Nb/Nb_2O_5/Ag$ (tested in-house and by [190]). More tested metal-insulator combinations are listed in [149].

ZrCuAlNi was suggested as an electrode due to its low roughness after growth, making it suitable for MIM devices[182,183]. However, it is not ideal due to formation of an interfacial compound at the interface with the insulator. This was highlighted by Grover & Moddel[168], who anticipated the importance of material selection for a good MIM and that arbitrary combinations of metals and oxides are not always feasible, as for example $Al/Al_2O_3/Pt$[173] displays a linear I-V and yet has been successfully shown to work in an antenna-coupled infrared detector. A study by Heiblum *et al.*[174] tested $Al/Al_2O_3/Al$, $Cr/Cr_2O_3/Au$, $Al/Al_2O_3/Ag$ and $Nb/Nb_2O_5/Au$ MIMs by exposing them to an IR laser and concluded their poor rectification properties due to the production of large thermal contributions when exposed to laser radiation. This is a very useful approach, which must be applied to other MIM candidates for solar applications. An alternative concept was previously described (Figure 16), consisting of a MIM diode based on a travelling-wave structure, which although having a symmetrical configuration ($Nb/Nb_2O_5/Nb$), outperforms planar type MIM diodes by three orders of magnitude at μm wavelength[134].

One of the most important findings in this field is a method for characterising the electrical properties of MIM devices by nanoindentation and later a bent wire technique suggested by Periasamy *et al.*[149,184–187]. The concept is based on a metal being deposited on a silicon substrate, after which a continuous oxide is grown and the second electrode is a bent wire of a desired material. This ensures a small contact area and the ability to characterise a point-contact type MIM junction. This work is very useful as it gives the chance to study various metal-insulator combinations before attempting to optimise planar thin-film diode devices. This technique was used to systematically study the influence the metal workfunction difference has on the electrical properties of MIM diodes (reference [149]). The structures analysed have a Nb bottom electrode, Nb_2O_5 insulator grown by anodic oxidation, and eight different metal 2 selections (Hf, Zr, Nb, Ti, Cu, Ag, Au, and Pt). The result was that the workfunction difference has a direct influence on rectification properties, with the asymmetry increasing sharply for $\Delta\Phi > k_BT$. This agrees with theoretical predictions and with the models by Simmons[191]. Conversely, nonlinearity FOM displayed no direct correlation with $\Delta\Phi$. The highest FOMs were observed for $Nb/Nb_2O_5/Pt$, which displayed 1500 asymmetry, 4 nonlinearity and 20 V^{-1} responsivity meeting the criteria outlined in Equations 18, 19 and 20. Recently, this work was continued through (reference [188]) the fabrication of 6400μm^2 planar devices with the same Nb/Nb_2O_5 base and six metal top contacts (Ag, Cu, Nb, Ni, Au and Pt). The work confirmed a direct linear relation between increasing asymmetry and increasing $\Delta\Phi$ with $Nb/Nb_2O_5/Pt$ showing optimal results with figure of merits of 7742 asymmetry, 4.7 nonlineary and 52.2 V^{-1} responsivity. These are to our knowledge the largest FOM values recorded and hence the most promising structure to be successfully integrated into a solar rectenna.

Periasamy *et al.*[185] also characterise other oxides with a Nb metal base and Pt bent wire and prove a hypothesis whereby the best rectification properties are expected from using dissimilar metals and low barrier heights at each metal-insulator interface in order to sooner promote the faster Fowler-Nordheim tunnelling in the forward direction. To achieve smaller barriers, high electron affinity insulators must be selected. Due to this reason, Nb_2O_5 and TiO_2 outperform Al_2O_3 and MgO. Another requirement for the insulator is a large dielectric constant since for small area devices the capacitance is reduced, however due to the requirements for tunneling the ultra-thin

thicknesses in turn increases capacitance, which has to be kept to a minimum as per Equation 2.

3. *Modelling approaches*

Modelling MIM devices is a significant challenge due to the many practical aspects of such ultra-thin layers. Interface roughness, pinholes and fabrication anomalies may all adversely influence device performance, and are difficult to account for reliably in the models. Nevertheless, modelling MIM performance is essential in order to confirm established experimental results and fully understand device operation as there is always room for improvement. Simmons[191] first proposed a model incorporating the WentzelKramersBrillouin (WKB) approximation, which is frequently used to reproduce MIM I-V curves. This model is very good at predicting forward characteristics, however it overestimates currents passing in reverse bias, leading to close to symmetric simulated I-V curves for dissimilar metal electrodes. Eliasson[165] developed a comprehensive model to predict rectification efficiency, tunneling transmission probability and estimation of device resistance in thin MIM diodes, with the exception of image forces and effective mass. Recently Hashem *et al.*[192] developed more accurate models to simulate the I-V curves and figures of merit of MIM devices. The model can be used for multiple insulator layers, whereas Simmons model is only limited to single insulator predictions. These models were shown to accurately trace experimental results reported by Choi *et al.*[193](on a polysilicon/SiO_2/polysilicon tunnel diode) whilst Simmons simulations predicted almost linear behaviour. With double insulator devices, the models are accurate in forward characteristics, however slightly underestimate the leakage currents in reverse bias. Recent work by Cowell III *et al.*[194] has developed a model to estimate the barrier height formation and tunneling effective mass of a MIM diode exhibiting Fowler-Nordheim tunneling, which can be used to help assess the impact of different processing conditions on the performance of the device.

4. *Fabrication and Characterisation*

Apart from material selection, the fabrication techniques used to manufacture MIM devices are also important for good device performance. Metal deposition is commonly deposited by sputtering or evaporation, techniques which are well understood, give high quality thin films and are industrially scalable. Controlled deposition of the ultra-thin insulator required for tunneling is a challenge. Anodic oxidation[149,185,188] of the first metal electrode is a good way of controlling growth of a high quality oxide, however, it is limited to only being able to grow the derivative oxide of the underlying metal. For research purposes, pulsed-laser deposition (PLD) supplemented by reflection high energy electron diffraction (RHHED)[195,196] can be used for monolayer controlled growth of oxide layers. This technique achieves ultra-thin continuous stoichiometric films of extremely high qualities, however PLD is not industrially scalable. Alternatively, reactive magnetron sputtering may be used, but there is some uncertainty over the stoichiometry and thickness control when depositing insulators. One of the most promising growth techniques is atomic layer deposition (ALD), which is a self-limiting deposition process allowing accurate thickness and stoichiometry control similar to RHEED PLD whilst maintaining a comparably high throughput and possibility of industrial scale processing.

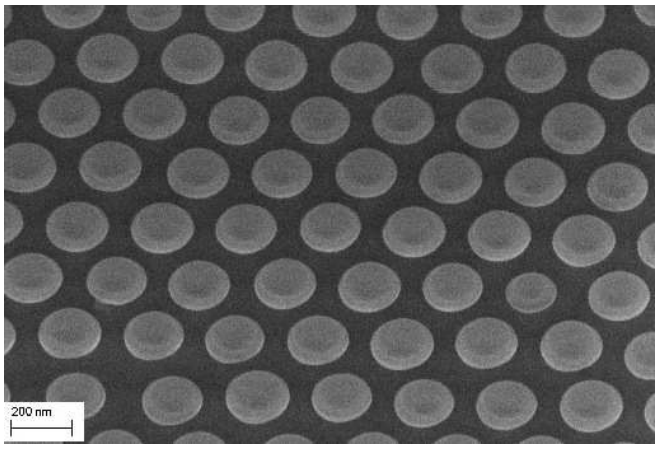

FIG. 25. SEM image showing an array of nanodiodes produced by growing an MIM multilayer, followed by colloidal lithography and Ar ion milling.

MIM device arrays of nanometer gap junctions over cm^2 areas[105], to create metal-vacuum-metal tunnel junctions which have similar properties to MIM diodes, also present a fabrication challenge. Common lithographic techniques have been used, but the economically viable techniques are limited to micron scale devices. However when the common lithographic techniques are supplemented by a recently optimised selective area atomic layer deposition technology[197], tunnel junctions of approximately 1 nm can be produced. This enables designs of antenna-rectifier coupled devices other than the more common planar types, which have great potential for success (see Miskovsky *et al.*[105]). Another challenge is fabricating arrays of nano-scaled lateral diodes to be of the same size scale as required for high frequency antennas. Bareib *et al.* have developed a highly efficient method for producing an array of several million nanodiodes by temperature-enhanced transfer priniting[198]. These devices were characterised with a conductive atomic force microscope setup (C-AFM), which showed good asymmetry in the I-V response of the nanodiodes. An alternative technique was developed by the authors, which involves the deposition in a vacuum of the full MIM layers, fol-

lowed by colloidal lithography and Ar ion milling to give an array of nanodiodes (Figure 25), and preliminary C-AFM confirms assymetry.

Current MIM tunnel diodes remain limited to rectify the infrared part of the solar spectrum due to their *RC* time constant. In a study to confirm the applicability of MIM diodes to solar rectennas, Grover & Moddel[168] conclude that efficient coupling between antenna and rectifier must be achieved, which implies that the rectifier impedance must be around 100 Ω as identified by Ma & Vandenbosch[95]. Results by Chin *et al.*[188] show promising improvement in MIM performance, although the structure has a limited current density passing through it due to the relatively *thick* oxide used (15nm). MIM diodes appear to be a good candidate for the rectifiers in a solar rectenna and the ever increasing success in modelling, characterisation and fabrication of these multi-layered devices have shown their potential.

C. Metal-Insulator-Insulator-Metal Diodes

The addition of another Insulating layer was proposed as it would benefit the rectification properties of the tunnel diode. Using Metal-Insulator-Insulator-Metal (MIIM) devices, the diode figure of merits are improved over the single insulator case. Asymmetry here can be achieved even with similar metal electrodes, as long as the electron affinity of both insulators is different. For the purpose of understanding the concept, we will qualitatively explain the simple case of Metal 1 and 2 being the same and Insulator 1 having a higher electron affinity than Insulator 2. At zero bias the band structure is the one depicted in Figure 26**a**. When a negative bias is applied to Metal 1 the bands shift as in Figure 26**b**, where the *effective* tunneling distance, d_t, is of the length of both insulator thicknesses and direct tunneling through both insulators is observed. Increasing the bias will eventually shift the bands to the situation in Figure 26**c** and the d_t is now shorter, however the electrons will fall into the Quantum Well (QW), from where the tunneling probability is reduced. By reversing the bias, initially tunneling will be direct through both insulators as in Figure 26**d**. Further increasing the bias will shorten the *effective* d_t and the tunneling is again direct but through only one of the insulators (Figure 26**e**). Further increasing the bias will further decrease d_t by promoting Fowler-Nordheim tunneling, whilst in the reverse case electrons will continue to tunnel into the QW. The origin of the asymmetry in the I-V curves of MIIM diodes with similar metal electrodes is due to the shift in dominant tunneling mechanism during forward and reverse bias and the formation of the Quantum Well due to using insulators of dissimilar electron affinity. By employing different metal electrons, the rectification properties of MIIM diodes can be further improved.

If the thickness of the lower electron affinity insulator is increased relative to the other insulator then the

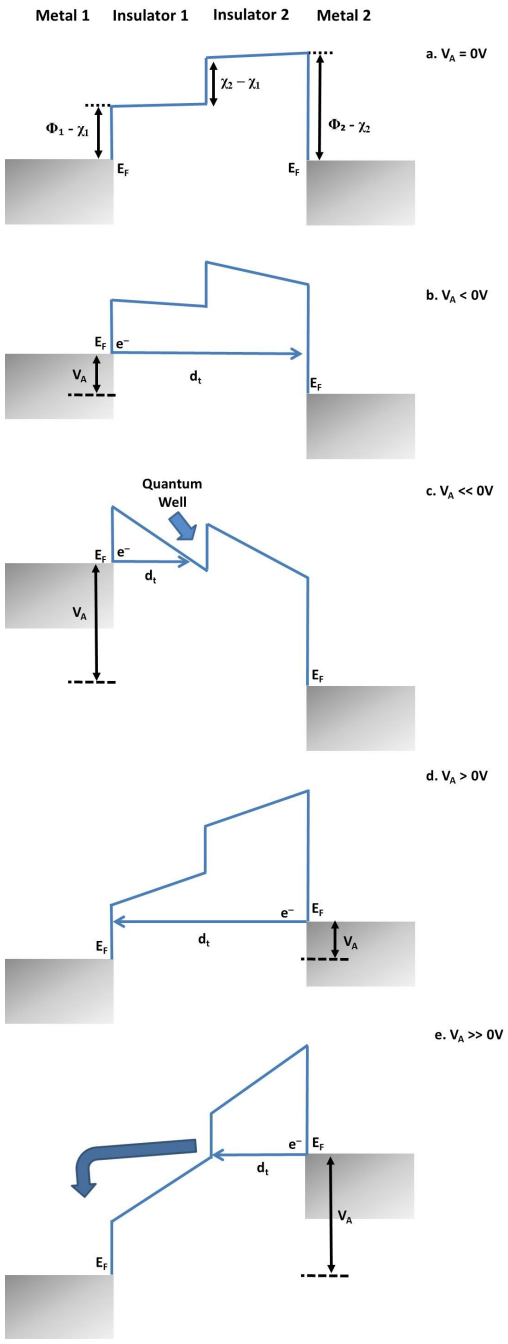

FIG. 26. Ideal band diagram for a MIIM diode with the same metal electrodes ($\Phi_1 = \Phi_2$) and two Insulators of equal thickness ($\chi_1 > \chi_2$). **a.** No bias situation ($V_A = 0V$); **b.** & **c.** Negative bias at Metal 1 (Reverse bias); **d.** & **e.** Positive bias at Metal 1 (Forward bias).

Quantum Well will be wide enough to form resonant energy levels[199,200]. These quantised resonant tunneling energy levels have electron transmission probabilities higher than that for adjacent energy levels. Therefore, they introduce significant asymmetry into the I-V curve[165] in the opposite direction to the one which would be observed

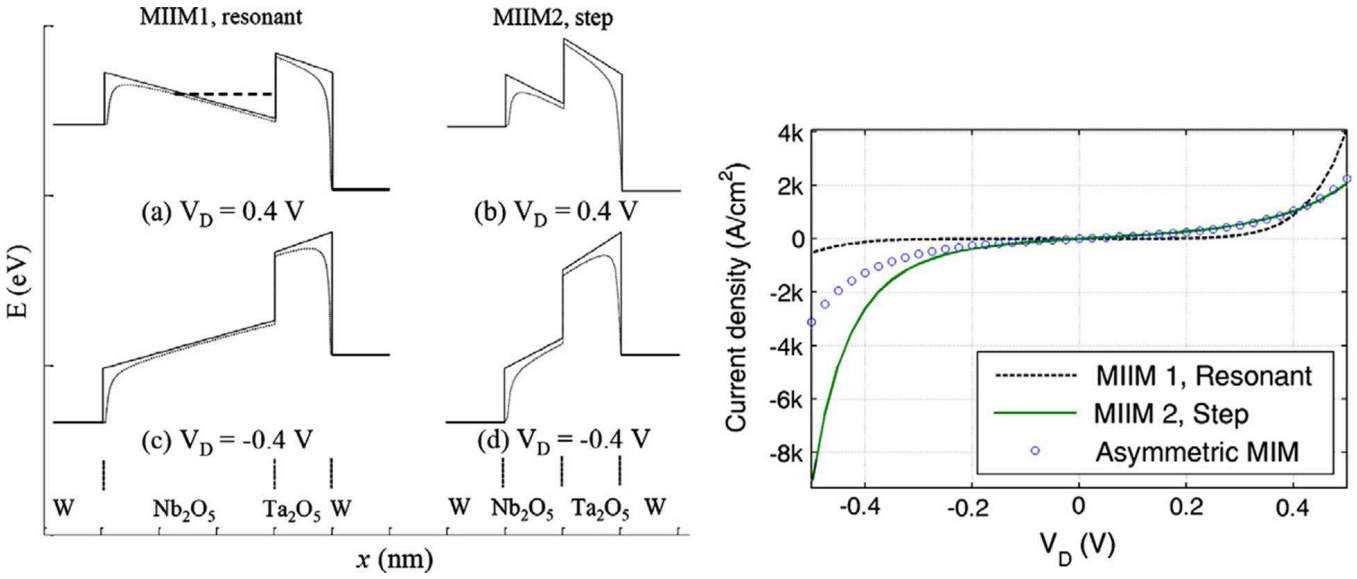

FIG. 27. Energy-band profiles for the W/Nb_2O_5/Ta_2O_5/W MIIM diodes discused by Grover & Moddel (left) under forward and reverse bias for the resonant (a and c) and step (b and d) diodes, and their corresponding I-V curves also compared to an Asymmetric MIM diode (right). Notice how current enhancement for the resonant diode is in forward bias, whilst that for the step diode is in reverse bias. (Reprinted from [200], ©2012 with permission from Elsevier)

if no resonant tunneling states are present (step MIIM), as discussed and shown by Grover & Moddel[200](see Figure 27). Adding extra insulators with resonant levels further increases rectification[201].

MIIM devices were first introduced for solar based applications by Phiar Corp. and patented by Eliasson and Moddel[202,203]. The performance of MIM and MIIM diodes as infra-red detectors was compared by Hegyi *et al.*[204] by simulating their responsivity. It was shown that the MIIM diode has 10 times higher responsivity than a derivative MIM diode. This was experimentally shown by Maraghechi *et al.*[205] who recorded up to 10 times better nonlinearity for a Cr/Al_2O_3/HfO_2/Cr diode compared to Cr/Al_2O_3/Cr and Cr/HfO_2/Cr diodes. A more applicable example, already referred to, is by Grover & Moddel[200], who compared their W/Nb_2O_5/Ta_2O_5/W structure to an MIM Diode with equivalent barriers as that of a W/Nb_2O_5 on one side and a W/Ta_2O_5 on the other side, again clearly showing improved response from the MIIM device (Figure 27).

Material selection for MIIM diodes has similar criteria to that of MIM, in that the structure has to exhibit one kind of tunneling in one direction and another tunneling mechanism in the other, thus promoting highly asymmetrical nonlinear I-V curves. A significant challenge is fabricating two ultra-thin, stoichiometric, high quality insulators. The appropriate technique for research purposes is RHEED PLD, although ideally ALD would be the better option in terms of prospects for industrial scaling. Alimardani *et al.*[206–208] have researched the use of ALD for depositing the double insulators required to produce efficient MIIM diodes. A recent patent[209] proposes the use of a triple-insulator barrier which adds further complexity to the fabrication process. The models created by Hashem *et al.*[192] will help identify which material systems to concentrate experimental efforts on. Improved performance from multiple barrier tunnel diodes is yet to be realised but the few reports that exist suggest significant potential. Integration with antenna for solar rectenna would also have to be researched as the addition of extra insulators may have deteriorating effect on the *RC* constant preventing these devices to be used at high frequencies although their rectification properties are clearly superior.

D. Geometric Diodes

Geometric diodes are devices that display rectification properties due to the difference in charge carrier transport probabilities from one electrode to another as a result of a geometric constraint. An increasingly popular concept patented by Moddel[210] consists of a thin film patterned into an inverse arrowhead[214] configuration shown in Figure 28(a). The electrodes are placed on top of the thin film at both ends of the structure. The width of the neck, d_{neck}, is of the same magnitude as the mean-free path length (MFPL) of the thin film material in order to obtain rectification[211–215]. Looking at Figure 28, an assumption is made by which charge carriers experience specular reflections at the boundaries[211]. At zero bias, the distribution of carriers is random and scattered across the left and right side of the structure and no net current flow is observed. Upon applying a bias, carriers gain

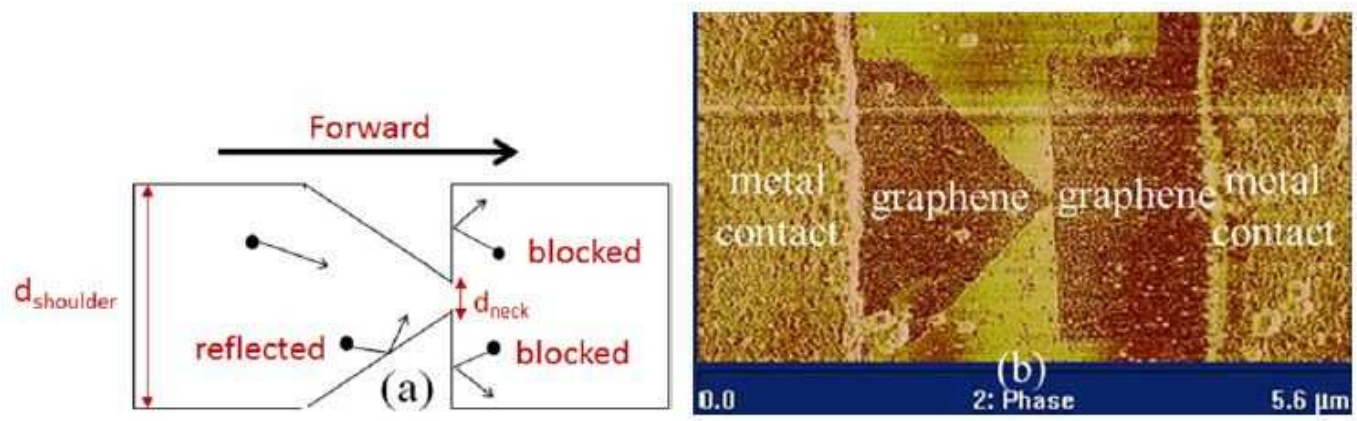

FIG. 28. (a) Schematic of a geometric diode theory of operation. (b) SEM image of a graphene geometric diode with metal contacts developed by Zhu *et al.* (Reprinted from [214], ©IOP Publishing. Reproduced by permission of IOP Publishing. All rights reserved.)

extra velocity in addition to their random thermal drift. For a plain non-constrained geometry, there will be equal net flow from one end to the other, which would define a completely linear I-V characteristic (Ohmic behaviour for perfect conductors). By applying a geometric constraint as the one shown in Figure 28, the motion of the carriers is influenced and favoured in one direction. Carriers moving from the left to right either pass directly through the neck or collide against the angled edge and eventually channel through the neck[211–214]. On the right hand side, most of the carriers are blocked and reflected away from the neck by the vertical blocks[211–214]. The probability of electrons passing from the left to the right is enhanced over the probability of them passing from right to left due to the *funnel-like* construction in the forward direction. The neck is required to be of the order of the MFPL, otherwise the probability of electrons passing in reverse direction will increase significantly and compromise the rectification performance.

Neck design is crucial to successful geometric diode performance and is therefore one of the governing factors in the material selection criteria for the device. The material would also have to withstand high current densities through the neck[214]. Metals are suitable to be used[168], however, their MFPL at room temperature is 10-30 nm[216], which complicates the patterning process and requires expensive techniques such as electron-beam lithography. Graphene has been identified as a suitable and promising alternative to be used in geometric diodes[210–212,214,215]. Advantages of graphene are that the MFPL can be tailored up to 1 μm[217] which means that common lithographic techniques can be used for device patterning and fabrication, resistances are low enough to be matched to the antenna impedance[214], and very low capacitance values of the order of 1 aF[212] lead to a low RC time constant of around 1.6 fs[211]. All these benefits identify their potential to be used at infrared and even optical frequency rectennas. Recent work on graphene geometric diodes has shown optical frequency rectification at 10.6 μm wavelength radiation (equivalent to 28 THz) matched with both a metal or a graphene bowtie antenna[214]. Further improvement in the design of these diodes could improve the performance of the device and potentially reach operational frequencies in the visible regime.

VI. RECTENNA DESIGN

Learning from the success in optimising the microwave rectenna, it can be seen that appropriate circuit design is of paramount importance for increasing the performance and efficiency of the device. The solar rectenna is an entirely different challenge as it is constrained by physical limitations and complex material selection criteria and trade-offs. Research is mainly concentrated towards manufacturing and characterisation of antenna and rectifier components, where novel technologies are emerging and great improvement has been seen over the years. Nevertheless, designing the best rectenna architecture is as equally important as optimising its two main constituents. There are many publication on antenna-coupled rectifiers used for optical[94,172,218–220] or infrared[108,134,173,176,177,221–223] sensing whereas, to the best of our knowledge, there is no antenna-rectifier device reported to harvest electromagnetic energy at the high solar frequencies. The gap between sensing and harvesting lies in how efficiently the incoming signal is converted and this is what limits the realisation of a true working solar rectenna. We review a few rectenna architectures proposed in the literature, which could result in future success.

Rectennas can either be designed as a planar device or have a geometrical constraint. The diode is used as a passive component with the sole purpose of rectifying the AC signal generated from the antenna. In this case

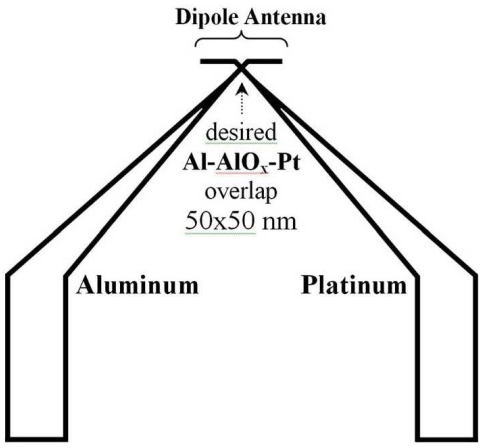

FIG. 29. Schematic diagram of an infrared detector proposed as a suitable design for a solar rectenna. The structure consists of a dipole antenna coupled to a metal-oxide-metal diode. (Reprinted from [173],©2009 with permission from American Vacuum Society)

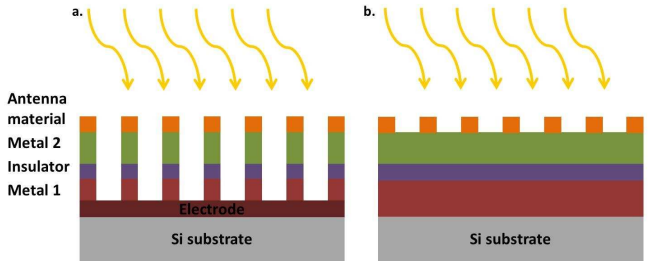

FIG. 30. Schematic diagram of a MIM-based planar-style rectenna architecture with **a.** rectenna array and **b.** antenna array with common rectifier.

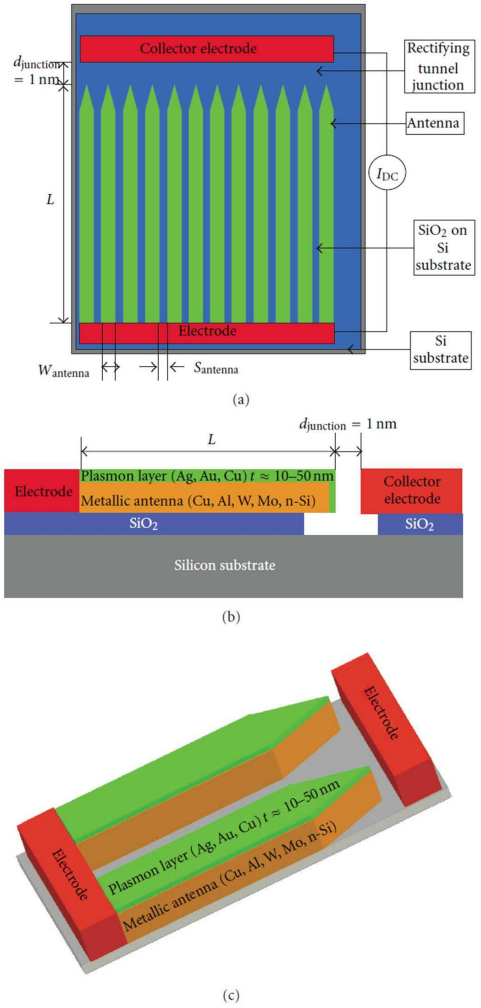

FIG. 31. Schematic diagram of the rectenna design proposed by Miskovsky *et al.* showing a top (**a**), side (**b**) and 3D (**c**) view. (Reprinted from [105], ©2012 N.M. Miskovsky *et al.*)

the requirements for the diode are emphasised on low resistance and capacitance (resulting in high cut-off frequency), and high I-V asymmetry at a small bias (implying low leakage current and low turn-on voltage). Both planar and geometric designs are suitable for employing this function of the diode. The following three architectures utilise the diodes ability to rectify current.

Figure 29 shows a dipole antenna coupled with a metal-oxide-metal diode based on an Al/AlO$_x$/Pt structure. This type of architecture is used for IR detection[134,173,177,221] and up to 28 THz[177] can be sensed with a Ni/NiO/Ni diode. Arrays of the structure can be fabricated either with e-beam lithography or a more economical nanotransfer printing process, which does not compromise the quality of the tunnel junction[221]. This design gives the benefit of producing small area diodes directly coupled to an antenna, which can be tailored to the length scale of the desired operating frequency. With appropriate material selection and improved impedance match, the cut-off frequency can be increased[177] and electromagnetic energy could potentially be harvested rather than just detected.

A common approach in rectenna design is a simple planar structure which is composed of antenna-rectifier arrays on a single chip (Figure 30a). The size of the individual Rectennas in the array depends on the aimed harvesting frequency, i.e. for the visible range the rectennas should be of nano-scale dimensions whereas micro-scale would suffice for the near-infrared. Alternatively, the device can be designed as having a common rectifier with an array of antennas (Figure 30b) – bowtie, patch, spiral, etc. The limitations are impedance matching and antenna and rectifier cut-off frequency. With the ever increasing technological advances and device modelling, the necessary goals are seen as realistic and feasible.

Finally, Figure 31 shows schematics of the approach proposed by Miskovsky *et al.*[105]. This design consists of an array of needle-like nanostructures, which have been designed to utilise a geometric constraint for the tunnel-

ing mechanism. At the tip of each nanostructure, a 1 nm junction is formed which can either form a MVM-type barrier or if an insulator is placed then the barrier will be an MIM-type. The fabrication of this type of structure is very complex, but was recently optimised by use of selective area atomic layer deposition – a technique developed and patented by Prof. Brian Willis of the University of Connecticuit[197]. Similar design can be implemented with the described Graphene geometric diode[214] where instead of leaving a tunnel junction, the tip of the needle-like nanostructure is brought close to the collector electrode. This rectenna design is very promising and yet to be optimised in terms of material selection, ease of manufacturing and reliability.

Another planar device design has already been introduced in Section V.A, whereby the rectifier is employed as an active component producing a current enhancement as a result of hot electron injection into the metal and crossing over a Schottky barrier, which is seen as a reverse current enhancement in an I-V plot (similar to what is observed during the photovoltaic effect). The design necessitates a diode with low barrier height, low leakage current and a suitable top metal to increase the amount of electrons excited to a higher energy state. It must be made clear that this device, although having components similar to a rectenna, is fundamentally different to the rectenna. With rectennas, the aim is to optimise the absorption by a load, whereas for hot-electron devices it is in optimising the absorption efficiency of the LSPR nanoparticle. A good review of hot-electron devices can be found in [224]. Efficiency estimates for rectennas prevail to those for hot-electron devices, although this remains an intriguing field for future development.

VII. SOCIO-ECONOMIC IMPACT & FUTURE WORK

The defining milestone for the rectenna was the Raytheon development of the microwave rectenna in the 1960s[13]. The ability to capture background radiation to demonstrate a working device, a helicopter no less, was a remarkable achievement. Concerns regarding energy security have led to a growing interest in harvesting solar radiation – a huge source of energy offering roughly 350 W/m^2. In 2011, the International Energy Agency predicted that solar energy technologies such as photovoltaic panels, solar water heaters and power stations built with solar collector mirrors could provide a third of the world's energy by 2060. Unsurprisingly, there was no mention of a rectenna device because there has never been a successful demonstration of a solar rectenna with even modest efficiencies of a few percent.

Solar cells made of silicon have already made a huge impact in a whole range of devices that use Si cells for power and they have also made a tremendous socio-economic impact in developing countries. It is worth recalling that around 1.2 billion people on our planet have no access to power and about 2.8 billion people use solid fuels (wood, charcoal, coal and dung) for cooking and heating[225]. Every year fumes and smoke from open cooking fires kill approximately 1.5 million people mostly women and children, from emphysema and other respiratory diseases and therefore a key responsibility for scientists and engineers is to carry out research to address such inequality. The advent of "plastic electronics" offers the possibility of inexpensive, flexible solar cells using semiconductors albeit at lower efficiencies of around 5-10%.

The rectenna is simply another means of collecting background radiation whether it be at optical frequencies in the solar spectrum (430-790 THz) or at IR frequencies (0.3-430 THz) meaning that cells could operate at night. The ultimate goal then of a rectenna is to augment current technologies and with current low frequency efficiencies of over 90%, the incentive to carry out research is obvious. However, the situation is not so simple and the key observation is that as the frequency of operation of the rectenna rises, the efficiency drops so that at solar frequencies most predicted efficiencies are a fraction of one percent. Here then is the first research challenge: to increase the efficiency to match at least the photovoltaic efficiencies offered by low cost plastic electronic solar PV (5-10%). Second, a rectenna device capable of operation at solar frequencies, which is a tremendous fabrication challenge with diodes of nanometre dimensions needed. Using expensive e-beam lithography techniques may demonstrate proof of principle but are too costly for commercialisation. Therefore, the engineering challenge is to find inexpensive fabrication techniques capable of scale-up and able to provide the efficiencies required.

So what are the prospects? There is little doubt that the task is hard – low cost fabrication of nano antennas and diodes of nm dimensions is non-trivial. But the rewards are significant. Perhaps most important is that, as noted in Section III, there is no scientific reason suggesting that a rectenna device operating at solar frequencies should not be capable of efficiencies in the region of 10% and ideally far greater. This then is the challenge and the potential reward is reason alone why there are several laboratories exploring such technologies. Let us be bold and assume success in this goal. What would be the implications? First of all, to have developed the technologies to fabricate the nanoantennas and diodes using scalable manufacturing processes would provide an extra option to the generation of energy. Second, success would not only impact upon a wide range of energy technologies but would be transferable such that it would also transform wide range of technologies associated with for example the Health, Transport and Infrastructure sectors. These sectors are fundamental to our socio-economic progress.

This review has identified many positive points on the development of the solar rectenna. Modelling the conversion efficiency has been attempted on many occasions, however as seen in Figure 8 the many different assumptions predict a wide spread of efficiency values. The sci-

entific community understands the operational theory of the device but further work is required to confirm and identify the major contributing loss and gain effects in order to accurately predict the efficiency. There is also lack of experimentally derived efficiency results at the high solar frequencies as research was until now mainly concentrated on developing new technology for nanofabrication and optimisation of antennas and diodes, modelling device response and theoretically estimating matching suitability. In the course of this review, new developments have been highlighted, which have given promising results for both the antenna and the rectifier. The technology is mature enough for it to be used to fabricate and characterise prototype rectenna devices. Testing prototypes will help identify further work, which needs to be carried out on the antenna and rectifier. A very useful methodology has been developed to characterise materials for MIM diodes and some experimental results of actual devices have been reported, identifying potential diodes which could be branded with having ideal I-V characteristics and far exceed the minimum figure of merit requirements. However, these diodes need to be characterised for their cut-off frequency and matching with an antenna. Work on the other rectifiers (MIIM and geometric diodes) is yet too undeveloped to have conclusive results, but provisional tests show great potential and they must be pursued and optimised.

In summary, despite the fact that Tesla suggested the transference of energy through electromagnetic radiation in the 1890s and despite the practical development of rectennas in the 1960-70s, the goal of solar rectification, or even large transfers of high-frequency power, has not yet been realised at the efficiencies we need. The two key issues that we need to address are to increase the efficiencies to useable levels, here suggested at 5-10%, and to develop low cost scalable manufacturing routes. The secondary goal is to increase the efficiency and here we can only speculate but we should harbour an ambition to obtain efficiencies exceeding that of the best solar cells tested to date of around 45%[226].

VIII. SUMMARY

In this article, the history and operation of the rectenna device have been reviewed from its initial concept as a receiver in microwave power transmission through to developments in pursuit of high-efficiency solar harvesting. Rectennas have been studied extensively for application in wireless energy transmission by use of man-made microwave energy with conversion efficiencies as high as 92%, which brings optimism to the goal of beating the Shockley-Queisser limit by achieving high-efficiency solar rectenna energy converters. Efficiency trends have been reviewed, where experimentally obtained efficiency values for rectenna devices have shown an overall decrease with an increase in operating frequency. Although there is no agreement on an optimised model for characterising the performance of a rectenna, all theoretical work suggests that to improve efficiency, three criteria must be met – 1) efficient absorption by the antenna at the desired frequency, 2) a high cut-off frequency diode to efficiently rectify the incoming signal and 3) impedance matching between antenna and rectifier to avoid signal transfer losses. Solving these three challenges would result in efficient device operation, which has the potential of operating at higher thermodynamic efficiency than current photovoltaic cells. Work on improving the performance of the antenna and rectifier has been reviewed showing substantial technological advancement and improved theoretical understanding of these two device components. Rectenna design also plays a major role in optimising the device and a few architectures have been discussed for their suitability. It was finally argued that the most essential future work required is to build rectenna devices based on optimised technology and design, as this is currently lacking in the literature. Attempts have been made on antenna-rectifier devices which have been aimed to work as IR or visible light sensors. Fabricating rectenna prototypes will enable scientists to characterise and assess their performance, giving them opportunity to identify new areas for necessary optimisation.

ACKNOWLEDGMENTS

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RF Energy Harvesting System And Rectennas

Riboy Cheriyan, Amsu Thejus. M, Jagadish Chandran G, Dr. K P Zacharia

Abstract —Recent advances in ultra-low power wireless communications and energy-harvesting technologies have made self-sustainable devices feasible. Typically, the major concern for these devices is battery life and replacement. Applying energy harvesting techniques to these devices can significantly extend battery life and sometimes even eliminate the need for a battery. This technique is widely applicable in low power circuits such as wireless sensors. One emerging wireless sensor networks (WSN) application is in agriculture sector, where sensor nodes are deployed in fields to monitor humidity, temperature and soil moisture. Energy supply to such sensors is an issue as they are typically powered by conventional batteries which have a limited lifespan. Cost is often prohibitive when replacing exhausted batteries since the sensor devices need to be unearthed. An attractive solution is to use radio frequency (RF) energy harvesting, in which the radiated RF energy from ambient is extracted and converted into usable energy to power up the sensors.

Index Terms — Harvesting energy, Patch antenna, Wireless sensors, Ambient energy , Schottky diode, Resonant circuit.

1. INTRODUCTION

Radio frequency energy is emitted by sources that generate high electromagnetic fields such as TV signals, wireless radio networks and cell phone towers, but through using a power generating circuit linked to a receiving antenna; this free flowing energy can be captured and converted into usable DC voltage. Most commonly used as an application for radio frequency identification tags in which the sensing device wirelessly sends a radio frequency to a harvesting device which supplies just enough power to send back identification information specific to the item of interest. The circuit systems which receive the detected radio frequency from the antenna are made on a fraction of a micrometer scale but can convert the propagated electromagnetic waves to low voltage DC power at distances up to 100 meters. Depending on concentration levels which can differ through the day, the power conversion circuit may be attached to a capacitor which can disperse a constant required voltage for the sensor and circuit when there isn't a sufficient supply of incoming energy.

We are being bombarded with energy waves every second of the day. Radio and television towers, satellites orbiting earth, and even the cellular phone antennas are constantly transmitting energy. What if there was a way we could harvest the energy that is being transmitted and use it as a source of power? If it could be possible to gather the energy and store it, we could potentially use it to power other circuits. In the case of the cellular phone, this power could be used to recharge a battery that is constantly being depleted. The potential exists for cellular phones, and even more complicated devices - i.e. pocket organizers, person digital assistants (PDAs), and even notebook computers - to become completely wireless.

Of course, right now this is all theoretical. There are many complications to be dealt with. The first major obstacle is that it is not a trivial problem to capture energy from the air. We will use a concept called energy harvesting. Energy harvesting is the idea of gathering transmitted energy and either using it to power a circuit or storing it for later use. The concept needs an efficient antenna along with a circuit capable of converting alternating-current (AC) voltage to direct-current (DC) voltage. The efficiency of an antenna, as being discussed here, is related to the shape and impedance of the antenna and the impedance of the circuit. If the two impedances aren't matched then there is reflection of the power back into the antenna meaning that the circuit was unable to receive all the available power. Matching of the impedances means that the impedance of the antenna is the complex conjugate of the impedance of the circuit.

This paper is considered to be one of the first steps towards what could become a standard circuit included in every wireless sensors, and quite possibly every electronic device made. A way to charge the battery of an electric circuit without plugging it into the wall would change the way people use wireless systems. However, this technology needs to be proven first. It was decided to begin the project with a cellular phone because of the relative simplicity of the battery system. Also, after we prove that the technology will work in the manner suggested, cellular phones would most likely be the first devices to have such circuitry implemented on a wide scale. This advancement coupled with a better overall wireless service can be expected to lead to the mainstream use of cell phones as people's only phones. This paper is an empirical study of whether or not this idea is feasible. This first step is to get an external wireless circuit to work with an existing phone by transmitting energy to the phone (battery) through the air.

2. POWER SOURCES

The main technological advancement that has allowed these harvesting devices to generate sufficient power is through the development of receivers which can sense broad ranges of frequencies, not just limited to television UHF signals, while capturing the highest concentration of generated waves. Since the

propagation of wave energy dissipates with distance from source, sensors located at far distances must extract energy at low power density which is related to the distance through $1/d^2$ [1]. A typical television antenna linked to the correct conversion circuit if located approximately 4 meters from the source of radio frequency generation of 677 mega Hertz and 960 kilowatts of effective radiation power is able to produce a detected voltage of 0.7 across an 8 kilo ohm load which is approximately 60 microwatts of harvested power and enough to power an LCD display thermometer[1].

Fig 1. Ambient power sources.

The output power of is limited by regulations, such as Federal Communication Commission (FCC), USA due to safety and health concern offered by EM radiations. The maximum theoretical power available for RF energy harvesting is 7.0 *u*W and 1.0 uW for 900 MHz and 2.4 GHz frequencies respectively for a free space distance of 40 meters.

A 1V/m electric field can yield power density of about 0.26uW/cm2. RF emissions from 800MHz range mobile equipments are significantly stronger signal than others, such as 2.4GHz range, which is mostly used by wireless LAN. A preferable range for RF energy harvesting is between 500 megahertz and 10 gigahertz, in which many different radio communication signals lies[4].

3. BASIC BLOCK DIAGRAM

Figure 2 shows the basic block diagram of RF energy harvesting system.

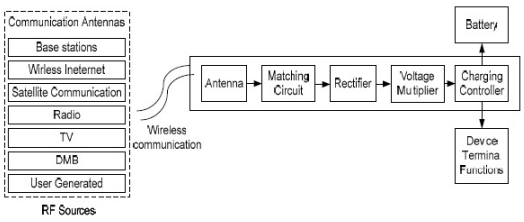

Fig. 2. Basic block diagram

Here we harvest energy from different sources such as mobile base stations, Wi-Fi, Radio waves etc. An antenna is used to capture signals, it may be a broadband antenna or separate antennas for different frequencies[2] .At high frequency we can use small antennas like patch antenna, but at lower frequencies we need a large antenna such as telescopic antenna in our FM radios[2]. Matching circuit consists of a variable capacitor for giving good impedance matching at the output of antenna at different frequencies. Rectifier is used to convert the ac at the output of antenna into DC. Here we prefer Schottky diodes because of its high frequency capability. Voltage multiplier consists of capacitor diode network which is used to increase the voltage level for charging the battery.

4. A SIMPLE GERMANIUM DIODE BASED ENERGY HARVESTING SYSTEM

Here a germanium diode based simple RF energy harvesting is shown in figure 3. Antenna used here is a stick antenna that is used in FM radios. The Gang capacitor and MW coil act as a matching circuit, resonating at a particular frequency.

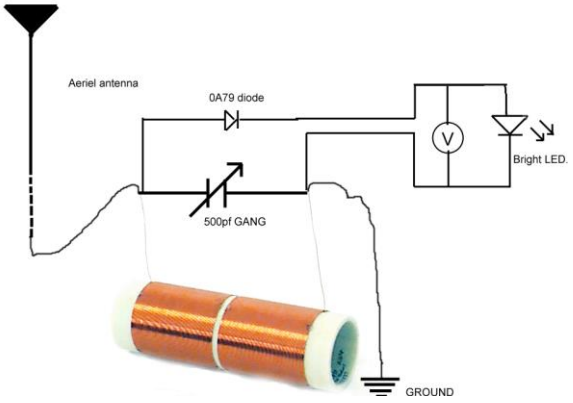

Fig. 3 OA79 Germanium diode based energy harvesting circuit.

OA79 Diode is used as rectifier because of its low turn on voltage. The gang capacitor tuned in such a way that the matching circuit resonates with the strong frequency that the antenna can capture. If the RF signal is strong, the circuit can drive a LED.

5. RF ENERGY HARVESTING SYSTEM BASED ON SCHOTTKY DIODE

Schottky diode offer low forward voltage and high switching speed, and consider as an ideal component for RF energy harvesting.

This circuit uses HSMS2820 schottky diode. RF signals can be captured using multiband antenna as

shown in figure. It is capable of receiving 900Mhz/1800MHz/1900MHz/2.4GHz. Small patch antennas can be used at higher frequencies.

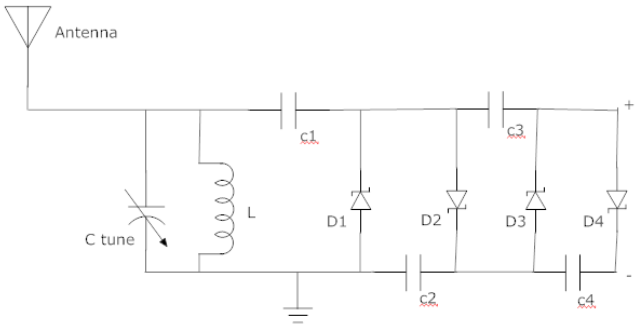

Fig. 4. Schottky diode based RF energy harvesting circuit with voltage multiplier.

Here combination of variable capacitor Ctune and inductor L forms the matching circuit for antenna at different frequencies. Using the Ctune in antenna matching circuit we can tune the antenna at its resonant frequency. The capacitor Ctune in tuning circuit can be varied using following formula to resonate with the antenna inductance L , were f is the frequency of operation.

$$2\pi f = \frac{1}{\sqrt{L_{antenna} \times C_{tune}}}$$

At higher frequencies such as 800 or 1900MHz the low values of inductors are difficult to construct especially at board level circuit design. But using the inductor along with capacitor at integrated circuit level design greatly improves the performance. Resonant frequency is also influenced by diode capacitance as it is related with reverse diode voltage and input voltage[4].

For converting Radio frequency to DC, a rectifier circuit is used .A charge pump circuit such as a voltage multiplier circuit can be used as rectifier. In figure there is a two stage voltage multiplier consists of two diodes and two capacitors in each stage. Here the diode used is a schottky diode (HSMS2820) because of its low forward voltage and high switching speed.

The voltage output of multiplier is twice input peak voltage minus the diode twice the diode threshold voltage.[4] The charge drained (Delta q) by the load current (I load) per period, where f is input signal frequency is

$$\Delta q = I_{load}/f$$

The circuit can be extended to n stages, producing the required output level. For a n stage multiplier the output capacitance can be calculated as

$$V_o = nV_{in} - \frac{n-1}{fC} I_{load}$$

From this, there should be a diode with very low turn on voltage and high operating frequency. Schottky diode has all this parameters so here a schottky diode is used.

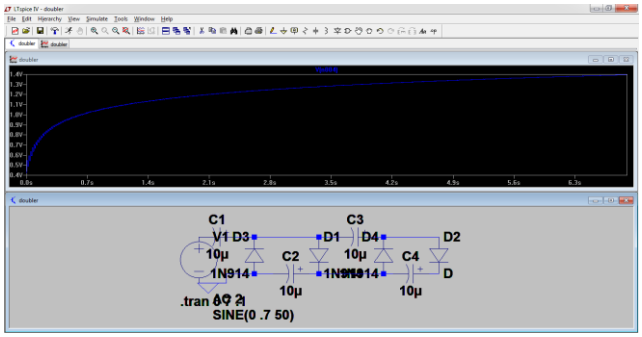

Fig. 5. Simulation result of two stage voltage multiplier.

Figure 5 shows the simulation result of a two stage multiplier with 0.7V input using LT Spice.

6. RECTENNAS

The rectenna termed as rectifying antenna, is combination of an antenna and a nonlinear rectifying element (Schottky diode, IMPATT diode…etc.) where the two elements are integrated into a single circuit. Since the rectenna is a receiving module collecting power from a radiation field, its sensitivity is defined by

$$S_R = \frac{output\ voltage}{power\ density\ at\ the\ antenna}$$

Division of this value with the effective antenna aperture leads to the normally used quantity for detectors.

$$S_D = \frac{output\ voltage}{input\ power}$$

This characterizes the nonlinear element with its matching.

A rectenna is useful as the receiving terminal of a power transmission system where dc power needs to be delivered to a load, through free space, for which physical transmission lines are not feasible. It is also suitable in applications where dc power needs to be distributed to more numbers of load elements that are spatially distributed. Such power distribution is achieved by the dispersive nature of microwave energy in space, eliminating the need for physical interconnects to individual load elements.

Figure 6 shows a rectenna for 899 Mhz. Here antenna consists of ainset fed rectangular patch with ground plane. It uses HSMS 2820 schottky diode as rectifying element

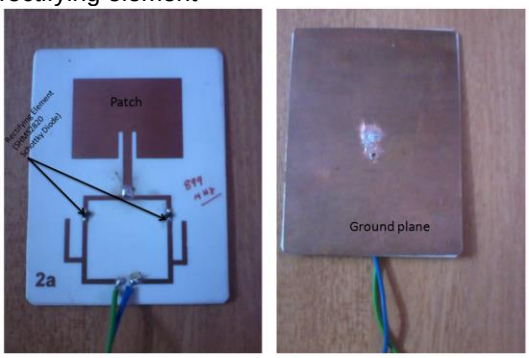

Fig. 6. Rectenna For 899 Mhz.

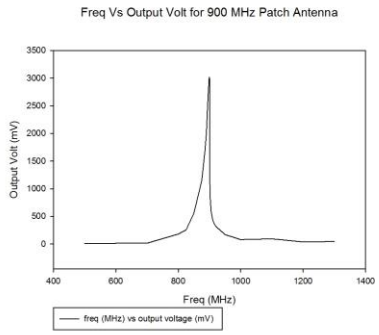

Fig. 7. Frequency Response of a 899 MHz Rectenna .

Figure 7 shows the frequency response of a 899Mhz rectenna. Here the peak is at 895 MHz. At this frequency we obtained an output dc voltage of 3.2V. In general it is difficult to predict how the rectenna system is optimized for the maximum conversion efficiency. However, there are several theoretical methods to overcome this problem.

The main objective with the design of rectenna is to obtain high conversion efficiency, and there are two approaches to achieve this goal. In first maximum power is collected and delivered to the rectifying diode, while in the second one harmonics generated by the diode are suppressed, which re-radiate from the antenna as power lost. In order to increase conversion efficiency by the first method, several broadband antennas, large antenna arrays and circularly polarized antennas have been designed and experimented. The broadband antenna enables relatively high RF power to be received from various sources while the antenna array can increase incident power delivered to the diode by enlarging antenna aperture and antenna gain. Antenna array is an effective means to increase the receiving power for rectification. However, a trade-off arises between the antenna size and the radiation gain. The circularly polarized antenna offers power reception with less polarization mismatch. However in second method, the rectenna consist of an LPF between the antenna and the diode, as well as an additional LPF on the dc output side of the diode. The main reason for the rise in the efficiency was the improvement in the diode and circuit construction for high input power levels.

7. CONCLUSION

In this paper, a study of various methods used for RFenergy harvesting has been made. It is found that we can harvest energy in micro watt range from ambient RF sources. Here the harvested power highly depends on the distance between the transmitter and RF harvesting system. By using an array of harvesting antennas we can harvest considerable amount of power.

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Antenna designs using matching circuits for 4G communicating devices

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UNIVERSITE DE NICE-SOPHIA ANTIPOLIS

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SCIENCES ET TECHNOLOGIES DE L'INFORMATION ET DE LA COMMUNICATION

ANTENNA DESIGNS USING MATCHING CIRCUITS FOR 4G COMMUNICATING DEVICES

THESE

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CA	Carrier Aggregation	**LTE**	Long Term Evolution
CE	Coupling Element	**MEMS**	Micro Electro-Mechanical Systems
EM	Electromagnetics	**MIMO**	Multiple Input Multiple Output
GSM	Global System for Mobile Telecommunications	**MN**	Matching Network
HB	High-Band	**PCB**	Printed Circuit Board
IFA	Inverted F Antenna	**PIFA**	Planar Inverted F Antenna
LB	Low-Band	**SMD**	Surface Mount Device
LDS	Laser Direct Structuring	**UMTS**	Universal Mobile Telecommunications System

ABSTRACT

With the introduction of the fourth generation (4G) of cellular mobile communications, there has been an increase in the frequency bands that a generic mobile phone should be able to operate. This requirement poses some challenges in the design of the transceiver modules of the device as well as the RF front-end and the antenna.

From the antenna design point of view, the newly added frequency bands especially LTE bands 12-13-17 which goes down to 700MHz creates the main design challenge: covering 700-960MHz (30% bandwidth) in the low-band, considering also backward compatibility with the former technologies like GSM850/900. Since the space reserved for the antenna is electrically small in this frequency range, special design techniques are necessary to obtain the required bandwidth for a generic mobile terminal antenna.

It should also be taken into account in the design phase, that the antenna needs to be appropriate for operation in the data-rate enhancement techniques like MIMO (multiple input multiple output) and carrier aggregation in LTE-Advanced. Furthermore, considering the real use cases of the mobile terminal, the effect of the user (hand, head and both) on the antenna performance should also be investigated.

In this thesis, several antenna designs for 4G coverage in mobile terminals are proposed. Different antenna topologies are evaluated consisting of tunable/passive antennas as well as single/multi-feed structures.

The "Introduction" chapter starts with a summary about the evolution of cellular communications starting from the first analog systems evolving to the fourth generation. General information about mobile phone antennas from the past decades is then presented ending with some examples of today's smartphone antennas. Electrically-small antennas are then presented briefly with their basic specifications and some state-of-the-art antennas for mobile terminals proposed in literature. The chapter concludes with a comparison between single-feed and multi-feed solutions with their advantages/disadvantages in LTE-Advanced technology.

Some tunable antennas, designed within the scope of the Artemos Project are presented in Chapter II. Firstly, two antenna designs using MEMS switches for band-switching between low-band (700-960MHz) and high-band (1.7-2.7GHz) are proposed, using antenna-parasitic strips for bandwidth enlargement. A circuit model is also proposed, modeling an antenna in combination with a grounded parasitic strip. A problem seen as radiation efficiency drop in these structures is also addressed with some solutions. Finally, a tunable antenna using tunable BST capacitors at the antenna feed is presented in this chapter.

A novel coupling element (CE) type antenna is proposed in Chapter III, which enables a more space-efficient design in comparison to plain CEs. First, the characteristics of the novel "hollow CE" is compared with a plain CE and two dual-feed antenna solutions are presented, using hollow CEs. As an application example, a MIMO structure using two hollow CEs is then given, using the neutralization line technique for higher isolation and lower correlation.

In Chapter IV, two single-feed antenna designs are proposed, manufactured using Laser Direct Structuring (LDS) technology. The first antenna can cover low-LTE/GSM/UMTS bands (700-960MHz and 1.7-2.2GHz) and the coverage of higher LTE and Wi-Fi bands (2.4-2.7GHz) is added in the second antenna. The effect of the user (hand and head) is investigated through measurements. An antenna resistant to detuning (due to the user) is then evaluated from these results as a future perspective.

A new use case of CE antennas is introduced in Chapter V, exploited to cover the 4G cellular communication bands in electronic eyewear devices, which are gaining popularity nowadays. First, a feasibility study is presented introducing the devices under design by several different companies. The bandwidth potential and SAR levels are evaluated according to different positioning of the CE in the eyewear device. Afterwards, three different antennas are proposed considering realistic casing and spectacles for the device.

Finally, a conclusion is done in Chapter VI, summarizing the general results from this thesis with some low-term and mid-term future perspectives.

CHAPTER I
INTRODUCTION

1. EVOLUTION OF CELLULAR COMMUNICATIONS

For more than thirty years, cellular communication has been under a substantial growth, according to the user demand. The first mobile communication systems were launched approximately around 1980s, giving the telephone users the freedom to make or accept calls, without the need for a fixed landline.

A cellular communication system is basically composed of three layers which are the core, the edge and the access subsystems [1]. Among these, the core is the nucleus of the network, that performs the handling of the voice calls, traffic control, user authentication and user profile (like charging the customer for the used services). The edge subsystem is an interface between the core and the access subsystems. The access subsystem (also called NodeB or Base Transceiver Station) is then responsible for establishing the air link with the mobile user equipment (UE). Different multiplexing schemes are used to serve a number of mobile users and duplexing techniques like FDD (Frequency Division Duplexing) and TDD (Time Division Duplexing) are used to establish a two-way communication (uplink and downlink) between the base station and mobile terminal. The land to be serviced is divided into some cells where these cells are served by base stations. The user can move between different cells and the handover is done automatically, transparent to the user.

The first mobile communications services (first generation or 1G) were launched in different parts of the world after 1979 [2]. These networks were based on analog systems for voice communications. The system used frequency modulation for modulating the voice signal and each user was assigned a specific frequency by Frequency Division Multiple Access (FDMA).

Approximately ten years later, the second generation of mobile communications (2G) was announced and the modulation scheme was changed from analog to digital, allowing also data transfer for services like Short Message Service (SMS). Multiple number of users were handled by using either TDMA (Time Division Multiple Access) or CDMA (Code Division Multiple Access). The most popular of the 2G standards is the GSM (Global System for Mobile Communications), which uses TDMA for multiplexing and FDD for duplex communication as uplink and downlink. GMSK (Gaussian Minimum Shift Keying) was chosen as the modulation scheme to provide better spectral efficiency. A main advantage of the GSM standard for the users was also the Subscriber Identity Module (SIM card), a card that carries the subscriber information and mobile number so that a user can change between different handset hardware by simply changing the SIM cards.

GPRS is a service launched on 2G systems, providing the user data services like MMS (Multimedia Messaging Service) and Wireless Application Protocol (WAP), making it being accepted as the 2.5G. GPRS was able to provide data rates up to 80kbit/s, which was increased later up to around 200kbit/s with the evolution to EDGE, called as the 2.75G.

With the release of the IMT-2000 standard by International Telecommunication Union, the downlink data rates of several Mbit/s started to be achieved. Systems like UMTS (W-CDMA), WiMax and CDMA2000 are examples of so-called 3G mobile communications, which introduced high-speed internet access, higher quality video and audio streaming. The most widespread radio interface is the W-CDMA using 2100MHz frequency for transmission over 5MHz bandwidth and FDD for duplexing. The downlink data rate was increased to 14Mbit/s and 28Mbit/s with the improvements called HSPA and HSPA+ (in the year 2008) respectively. A further improvement on the 3G systems towards 4G was the LTE (Long Term Evolution, 3.9G), targeting downlink data rates up to 100Mbit/s. LTE specification allows the usage of QPSK, 16QAM and 64QAM modulation types and channel bandwidth of up to 20MHz, with FDD and TDD for duplexing schemes. Additional usage of MIMO

(Multiple Input Multiple Output) can increase the maximum data rate from 100Mbit/s to 326Mbit/s. The access scheme used for the downlink is OFDMA (Orthogonal Frequency Division Multiple Access) for downlink and SC-FDMA (Single Carrier Frequency Division Multiple Access) for uplink.

2. GENERAL INFORMATION ON 4G TECHNOLOGIES

The successor of the 3G communication systems is the LTE-Advanced (LTE-A) which is accepted to be the fourth generation (4G) of the cellular communication systems. With this technology, the modulation schemes of QPSK, 16QAM and 64QAM are used to obtain peak data rates of up to 1Gbit/s for a low-mobility user in downlink (100Mbit/s for high mobility) and 500Mbit/s for uplink. OFDMA is used for multiple access in the downlink whereas SC-FDMA is used for uplink. The exploitation of multi-antenna systems (both for receive diversity or MIMO) is also available to reach the peak data rates. The available frequency bands for the 4G systems are shown in Figure 1, including also the 3G bands. A 4G mobile terminal should also be able to operate in the former 3G bands for backward compatibility since there is a risk of the lack a 4G base station everywhere the user may be located.

Figure 1 Frequency Allocation for 4G Systems

The LTE/4G frequency bands are presented in Table 1 and Table 2 for FDD LTE and TDD LTE respectively. It can be seen that the channel bandwidths allocated to each band is variable between 1.4MHz and 20MHz. Generally for the lower frequency bands (like bands 12 and 13) the maximum allowable channel bandwidth is up to 10MHz whereas for higher frequency bands (like band 3 or 10), bandwidths up to 20MHz is possible per channel. The multiple access scheme used in the 4G systems is OFDMA (Orthogonal Frequency Division Multiple Access), making the system more robust to fading, allowing higher data rates.

Another novelty introduced with LTE-A is the carrier aggregation, which is basically the utilization of more than one available channel for a single user. This way, the maximum achievable data rate offered to the customer can be increased. Three types of carrier aggregation are available for LTE-A as shown in Figure 2. The first type marked as (a) uses two adjacent frequency channels for a user which is called contiguous intra-band carrier aggregation. In the second type, two channels in the same band are aggregated, which are separated by other channels, making it non-contiguous intra-band carrier aggregation. Finally in the final case, two channels from two different bands are used together, which is called inter-band carrier aggregation. For example for a service provider with a license for transmitting in LTE Band 2 and Band 13, it is possible to provide data to a user over these two channels (20MHz in Band 2 and 10MHz in Band 13, making a total of 30MHz maximum bandwidth) simultaneously using inter-band carrier aggregation. A maximum bandwidth allocation of 100MHz per user is possible through carrier aggregation in LTE-A.

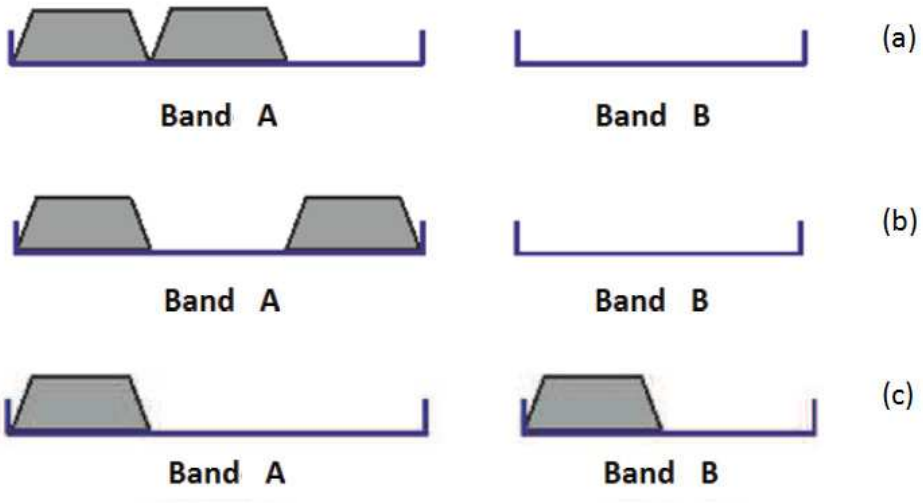

Figure 2 Carrier Aggregation Methods in LTE-A

Table 1 LTE FDD Frequency Bands

LTE Band	Area	DL Frequency	UL Frequency	Bandwidth (MHz)
1	All	2110-2170MHz	1920-1980MHz	5,10,15,20
2	NAR	1930-1990MHz	1850-1910MHz	1.4,3,5,10,15,20
3	All	1805-1880MHz	1710-1785MHz	1.4,3,5,10,15,20
4	NAR	2110-2155MHz	1710-1755MHz	1.4,3,5,10,15,20
5	NAR	869-894MHz	824-849MHz	1.4,3,5,10,15,20
6	APAC	875-885MHz	830-840MHz	5,10
7	EMEA	2620-2690MHz	2500-2570MHz	5,10,15,20
8	All	925-960MHz	880-915MHz	1.4,3,5,10
9	APAC	1845-1880MHz	1750-1785MHz	5,10,15,20
10	NAR	2110-2170MHz	1710-1770MHz	5,10,15,20
11	Japan	1476-1496MHz	1428-1448MHz	5,10
12	NAR	729-746MHz	699-716MHz	1.4,3,5,10
13	NAR	746-756MHz	777-787MHz	5,10
14	NAR	758-768MHz	788-798MHz	5,10
17	NAR	734-746MHz	704-716MHz	5,10
18	Japan	860-875MHz	815-830MHz	5,10,15
19	Japan	875-890MHz	830-845MHz	5,10,15
20	EMEA	791-821MHz	832-862MHz	5,10,15,20
21	Japan	1496-1511MHz	1448-1463MHz	5,10,15
22		3510-3590MHz	3410-3490MHz	5,10,15,20
23	NAR	2180-2200MHz	2000-2020MHz	1.4,3,5,10,15,20
24	NAR	1525-1559MHz	1626.5-1660.5MHz	5,10
25	NAR	1930-1995MHz	1850-1915MHz	1.4,3,5,10,15,20
26	NAR	859-894MHz	814-849MHz	1.4,3,5,10,15
27	NAR	852-869MHz	807-824MHz	1.4,3,5,10
28	APAC	758-803MHz	703-748MHz	3,5,10,15,20
29	NAR	717-728MHz	-	3,5,10
30	NAR	2350-2360MHz	2305-2315MHz	5,10
31	CALA	462.5-467.5MHz	452.5-457.5MHz	1.4,3,5

Table 2 LTE TDD Frequency Bands

LTE Band	Area	Frequency	Bandwidth (MHz)
33		1900-1920MHz	5,10,15,20
34	EMEA	2010-2025MHz	5,10,15
35	NAR	1850-1910MHz	1.4,3,5,10,15,20
36	NAR	1930-1990MHz	1.4,3,5,10,15,20
37	NAR	1910-1930MHz	5,10,15,20
38	China	2570-2620MHz	5,10,15,20
39	China	1880-1920MHz	5,10,15,20
40	China	2300-2400MHz	5,10,15,20
41	All	2496-2690MHz	5,10,15,20
42		3400-3600MHz	5,10,15,20
43		3600-3800MHz	5,10,15,20
44	APAC	703-803MHz	3,5,10,15,20

3. EVOLUTION OF MOBILE PHONE ANTENNAS

The evolution of the cellular mobile communications systems have also changed the characteristics of mobile phone antennas. This evolution was sometimes driven by technical requirements like newly added frequency bands as well as user demand about aesthetical properties. Detailed analysis about the evolution of mobile terminal antennas and possible techniques are presented in [3] and [4].

Older mobile phones had an external antenna on the top of the chassis through the end of 1990s. These antennas were generally either helix type, monopole (whip) type or a combination of both for dual-band operation. The length of the whip antenna was arranged as a quarter of the operating wavelength, forming an even better configuration if the metallic chassis of the phone or the PCB was also a quarter wavelength long, therefore having a full structure like a dipole. To reduce the length of the whip antenna, a good alternative was the helical (or helix) antenna, which can be thought of the convoluted version of the whip antenna (Figure 3). To obtain a multiband structure, it was possible to use a helix antenna with a whip passing through the center axis (Figure 4), two helix antennas with different radii sharing a common axis or two helical antennas with different pitches. Despite having very good efficiency performance, the external antennas had a drawback of having high Specific Absorption Rate (SAR).

Figure 3 Examples of External Antenna for Mobile Phones

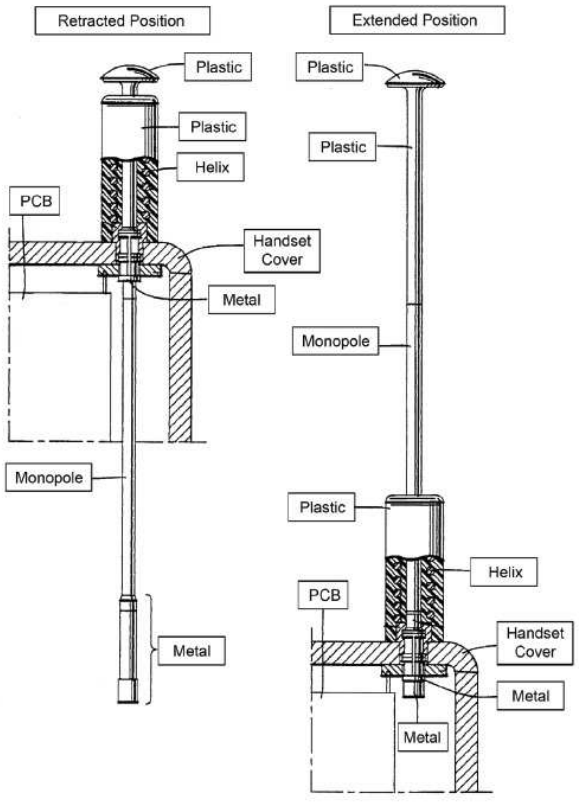

Figure 4 A Monopole & Helix Antenna for Mobile Phones, [3]

Through the end of 1990s, mobile phones having internal antennas started to be released on the market. Nokia released two models in 1998 and 1999 (8810 and 3210), which had internal antennas (Figure 5). One of the major internal antenna type, which was extensively used in mobile phones is the Planar Inverted F Antenna (PIFA), which can be thought of a patch antenna with a connection to the ground plane from some physical point of the antenna[5-6]. Multiband behavior was obtained easily by creating some slots in the rectangular plate to alter the current paths (Figure 6). Using some

parasitic elements either connected to ground plane on one end or left floating in both ends was also an alternative for multiband operation.

Figure 5 Mobile Phone Models Switching From External to Internal Antennas

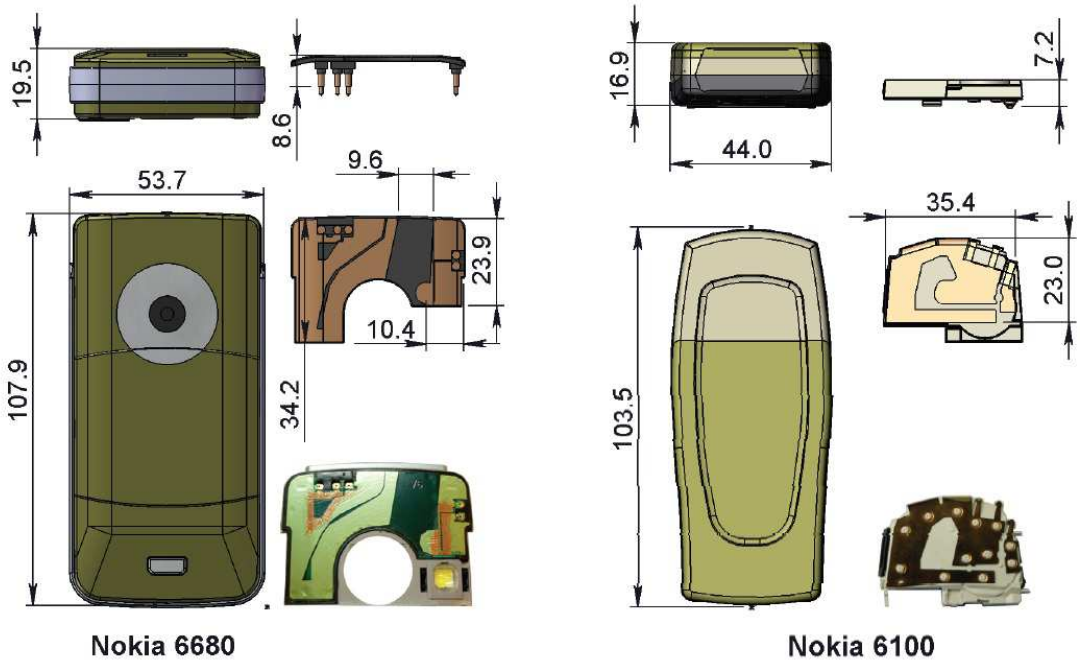

Figure 6 Examples of PIFA Antennas Used in Mobile Phones [3]

Another major internal antenna type used in mobile phones was the planar monopole antennas, which do not have a direct connection to the system ground plane. This type of antennas may require a ground clearance region in the vicinity (generally on one edge of the PCB) for acceptable performance. Techniques like adding parasitic elements or adding parallel branches to the radiating element can be used for multiband operation. This type of antenna was not used in the mobile terminals until recently due to the high SAR values exceeding the limits. The opportunity to place the antenna at the bottom of the mobile phone enabled their usage, since the SAR values were now suitable due to longer distance between the antenna and head when the antenna is on the bottom of the mobile terminal.

Throughout the evolution of the mobile phone antennas, it was generally needed to minimize the volume/area occupied by the antenna. Techniques like etching slots or meandering were used to alter the currents on the antenna element and/or the PCB, called inductive loading. Similarly, capacitive loading was also used for this purpose, this time to manipulate the electrical near fields (like bending the antenna closer to the ground plane). Dielectric loading was also a common method by covering the vicinity of the radiating structure with a relatively high permittivity material to decrease the resonance frequency.

As new standards and new frequency bands were introduced to the mobile communication systems, antennas with wider operating frequency bandwidth in approximately the same volume were needed. Parasitic elements (either grounded or floating) which are excited capacitively by the main antenna, creating additional resonances or changing the frequencies of existing resonances could be used for bandwidth enhancement. Another common technique is also the usage of matching networks, which can either be distributed along the antenna element (for example an inductor on the grounding point of a PIFA) or introduced directly at the feed for arranging the antenna input impedance in a desired way. The matching network can be either composed of completely passive elements (like SMD capacitors and inductors), completely active elements (like a tunable varactor at the antenna feed to tune the impedance) or a combination of both.

4. EXAMPLES OF ANTENNAS IN MODERN SMARTPHONES

This section aims to presents some examples of antennas and their placement in modern smartphones. The first example is the iPhone 4GS, which uses "coupling element" type antennas, whose general properties will be detailed later in this chapter. In iPhone 4GS, the metallic portions of the phone casing (referred to as bezel) are used as radiating structures. For this purpose the bezel is split into two, longer one forming the GSM/UMTS antenna and the shorter other forming the Bluetooth/Wi-Fi/GPS antenna. However, this structure had the problem of being very sensitive to user's hand, especially when the finger is placed between the gap separating the two metallic structures.

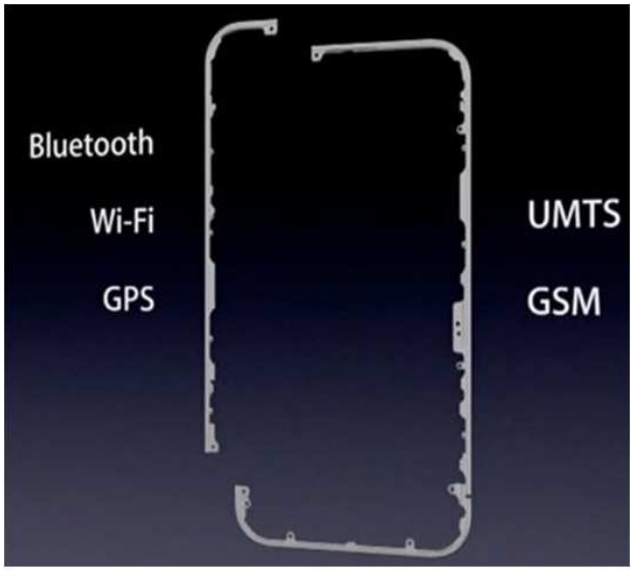

Figure 7 Antenna Placement of iPhone 4GS

In iPhone 5, the cellular antenna seen in Figure 8 was placed at the bottom of the phone and another antenna responsible for Wi-Fi operation was used at the top.

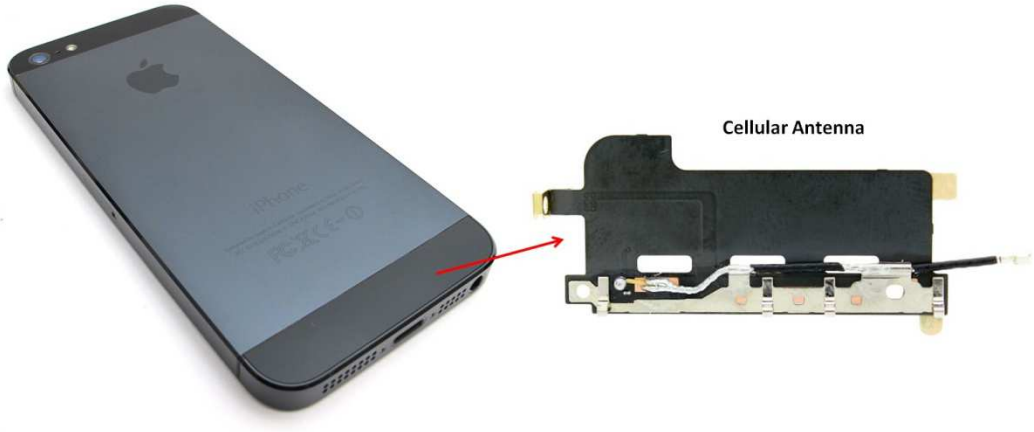

Figure 8 Antenna Placement of iPhone 5

A patent from Apple. Inc about a mobile phone antenna using again the metallic portions (bezel) of the phone casing as the radiating element can be seen in [7]. The antenna proposed in the patent is a loop antenna structure which is fed directly from the gap (shown as 18 in Figure 9), to minimize the fields across it and thus minimize any user effects due to the finger touching the gap (like the problem in iPhone 4GS). For impedance matching, a shunt inductance (formed by a metallic strip shown as 70 in the figure) and a series capacitance (shown as 100) is used at the antenna feed.

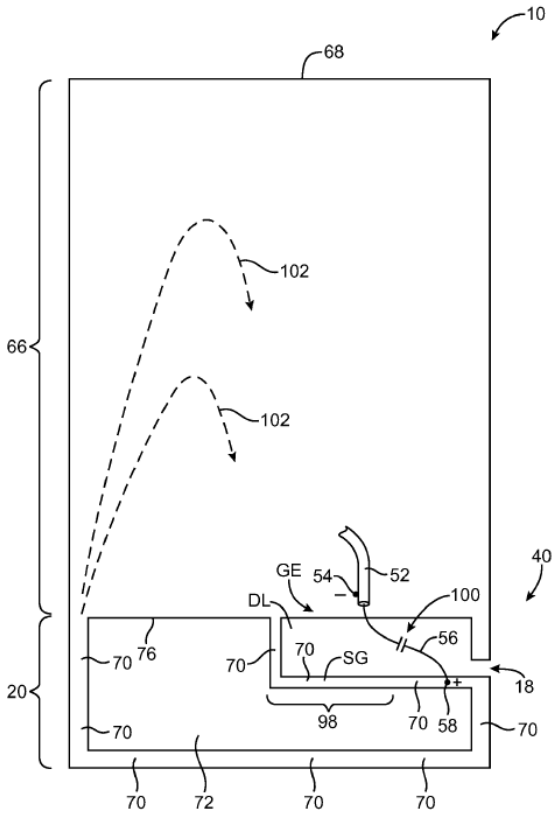

Figure 9 Antenna Structure Proposed in [7]

Another example is HTC Desire S as seen in Figure 10. This phone has three antennas realized using MID technology on plastic. The Bluetooth/Wi-Fi and GPS antennas are situated at the top of the phone whereas the IFA type pentaband antenna for GSM/UMTS is at the bottom.

Figure 10 Antenna Placement of HTC Desire S

The final example is Samsung Galaxy SII, which houses three antenna structures manufactured using MID technology, similarly as in HTC Desire S. In this phone, the GSM and Bluetooth/Wi-Fi antennas are located at the bottom and the GPS antenna is located at the top, just over the camera.

Figure 11 Antenna Placement of Samsung Galaxy SII

5. THEORY OF SMALL ANTENNAS

The popularity of internal antennas for mobile phones has gained a great popularity by the users after their introduction. This is why, the external antenna structures have quickly disappeared from the mobile terminal market. Since the frequency bands used for both GSM850/900, as well as the newly added frequency bands for LTE (like bands 12-13) go down to 700MHz, the free space wavelength rise up to 43cm for cellular communication bands. Considering the space that is available for the antenna in a mobile phone (generally lower than 20mm X 50mm X 7mm), it is evident that the antenna will be electrically small in the target frequency bands. It is also worthwhile to note that the space reserved for the antenna is subject to further decrease in smart phones having larger and larger touch screens and more and more electronic components to be placed like USB connector, camera, battery etc. The main drawback of an electrically small antenna is the low real part and high imaginary part of impedance, which makes it naturally narrowband. Electrically small antennas have been largely investigated in the literature [8-13].

A fundamental and intrinsic parameter of an electrically small antenna is its Quality Factor (Q), which is defined as the ratio of 2π times the energy stored in the near fields excited by the antenna to the energy radiated plus dissipated per cycle. The potential bandwidth of an antenna is inversely proportional to its quality factor, which results in low bandwidth potential for a high-Q antenna. The lowest obtainable radiation quality factor can be calculated from Equation-1 which is known as the Chu limit for electrically small antennas.

$$Q_r = \frac{1}{(ka)^3} + \frac{1}{(ka)} \qquad \text{(Equation-1)}$$

where k is the wavenumber (2π/λ) and a is the radius of the smallest sphere that can enclose the antenna. It is necessary to note that this value is an optimistic theoretical value considering a linearly polarized antenna radiating only in one (TE or TM) mode.

$$FBW = \frac{1}{Q}\sqrt{\frac{(TS-1)(S-T)}{S}} \qquad \text{(Equation-2)}$$

The obtainable frequency bandwidth can be calculated from Equation-2 as also presented in [14], where S is the maximum target VSWR and T is the coupling coefficient as given in [15]. For a value of T=1, Equation 2 reduces to:

$$BW_{cc} = \frac{1}{Q}\frac{(S-1)}{\sqrt{S}} \qquad \text{(Equation-3)}$$

This situation is called the "critically coupled" case, where the input impedance of the antenna is matched exactly to Z_0 (generally 50Ω) at a single frequency point (red curve in Figure 12). An alternative to get higher impedance bandwidth is to get optimal overcoupling, where the antenna is not matched exactly to Z_0, but matched in a wider bandwidth with a higher $|s_{11}|$ (for example with -7dB instead of -15dB), still under a predetermined target reflection coefficient level (blue curve in Figure 12). Some examples of using a matching network to obtain optimally overcoupled antennas are presented in [16-17].

For the calculation of the Q-factor of an antenna from its input impedance, Equation-4 was proposed in [13], which proposes a fast and reliable calculation method.

$$Q(\omega) = \frac{\omega}{2R(\omega)} \sqrt{[R'(\omega)]^2 + \left[X'(\omega) + \frac{|X(\omega)|}{\omega}\right]^2} \qquad \text{(Equation-4)}$$

where ω represents the angular frequency point, R(ω) and X(ω) stands for the real and imaginary parts of the impedance at angular frequency ω, and R'(ω) and X'(ω) are the first derivates of R(ω) and X(ω) according to the frequency. This formula is valid for a single resonant antenna close to its resonance frequency but it has proven reliable performance so far.

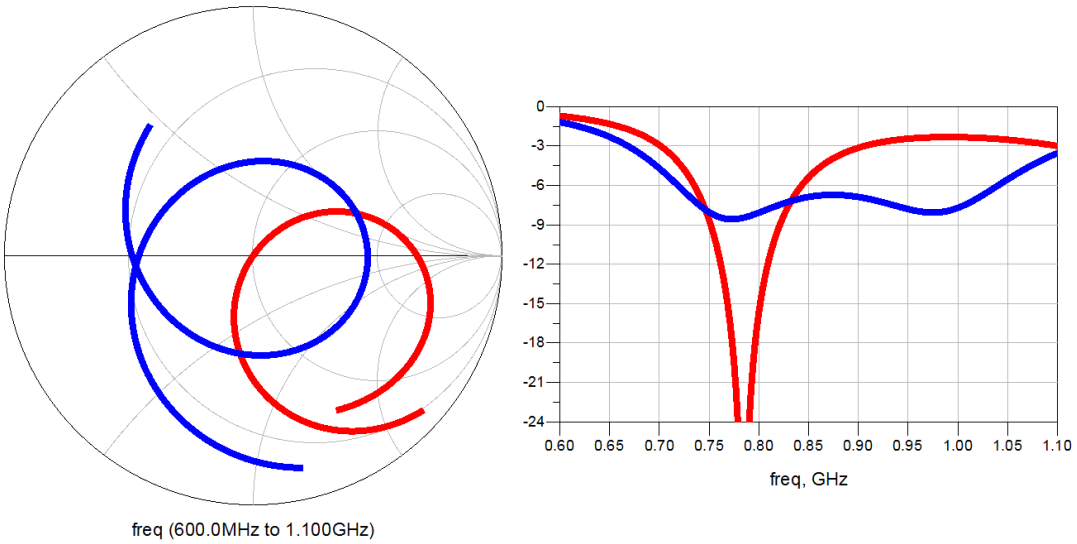

Figure 12 Critical Coupling vs Optimal Overcoupling

6. COUPLING ELEMENT ANTENNAS

As stated also in [18-21], the ground plane of the printed circuit board (PCB) in a mobile terminal plays the major role for radiation especially in the lower-frequency bands, where 90% of the total radiation comes from the PCB at 900MHz for instance[21]. The role of the ground plane in the lower frequency bands occurs regardless of the type of the antenna used, which is evident from the dipole-type vertically-polarized radiation pattern observed at low frequencies. The ground plane is also effective on the radiation in the higher bands being responsible for up to approximately 50% of the total radiation at 1800MHz.

Considering this fact, coupling element (CE) antennas were widely proposed mostly in the literature, to cover the cellular communication frequency bands in a mobile phone. Coupling element antennas are inherently non-resonant antennas, designed with the purpose of exciting the proper currents (chassis wavemodes) on the system ground plane [20-22]. In this antenna-scheme, the currents are induced on the PCB by capacitive excitation and the excitation is stronger close to the resonance frequencies of the PCB (which depends on its dimensions). For a 100mm ground plane length, the first chassis wavemode (longitudinal mode) occurs around 1.1GHz if a 10mm distance is set between the CE and this PCB. The feeding mechanism of the CE is similar to the one of a monopole, with no inductive contact to the PCB ground plane. In this way, the antenna-structure exhibits a low quality-factor and thus high bandwidth potential (BP) is obtained. Figure 13, which is taken from [14] explains the general behavior of a CE placed on a 100mm X 40mm ground plane without any ground clearance (CE placed directly on the ground plane). The major axis wavemodes

(first and second) are shown in the figure, occurring at 1.17GHz and 2.61GHz respectively. These wavemode resonances affect the input impedance of the antenna by slowing down the change of the impedance around these frequency points due to the low-Q property of the relatively wide ground plane, as presented in the smith chart. This effect can be best observed in the plotted quality factor, having some minima in these frequencies, corresponding to high bandwidth potentials. The antenna shall then be matched to the desired target frequency band using an appropriate matching network at the feed.

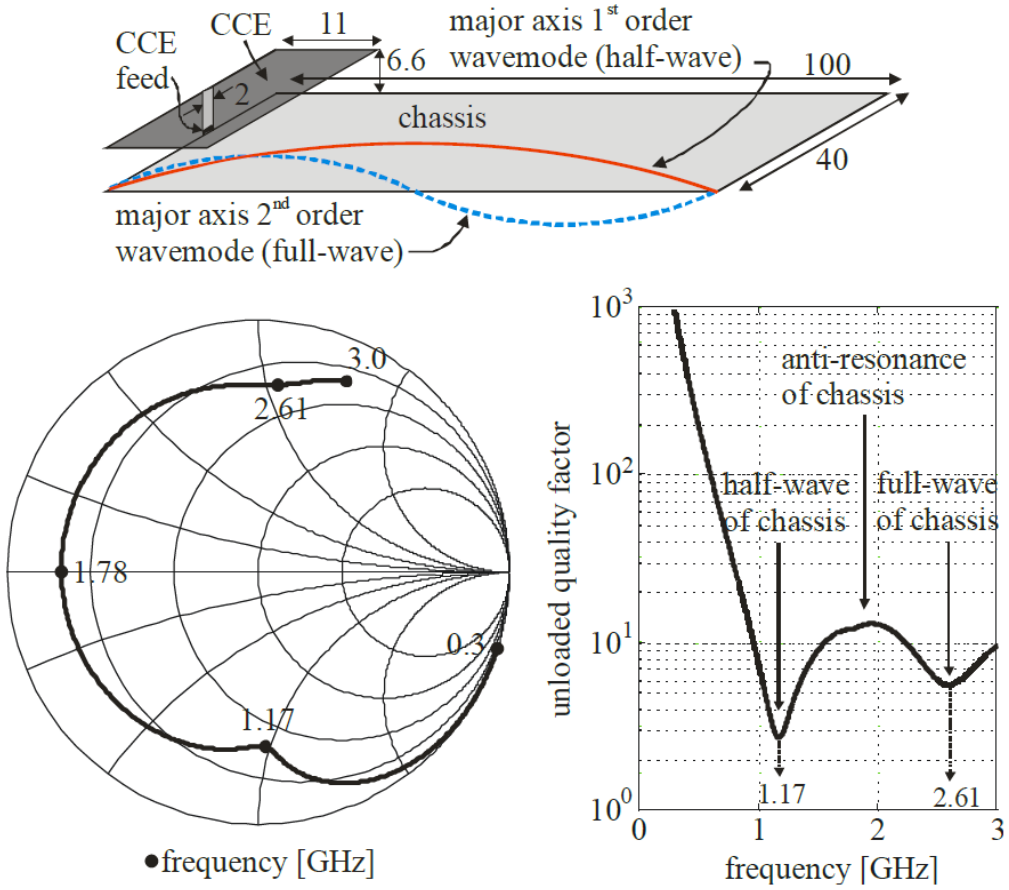

Figure 13 Summary of the Coupling Element Antenna Behavior

Despite the high obtainable bandwidth potential, the main drawback of this approach is the need for a tunable or a complex MN (complex in terms of design or large number of SMD components) if dual-band coverage is needed which increase the losses due to internal resistances of those SMD components. Another disadvantage is the necessity to keep the region between the CE and the ground plane clear from any metal components (ground clearance) that can distort the E-field distribution in this area and thus degrade the BP. Some CE designs targeting only the GSM/DCS/PCS/UMTS bands have been already presented in [23-24]. In [23], two CEs were placed on one of the short edges of the PCB, one responsible for the coverage of GSM frequency band (824-960MHz) and the other for DCS/PCS band (1.71-1.99GHz), with their associated MNs at each feed (Figure 14). To obtain a single feed topology in the final design, special attention was paid on the MNs, to get a high impedance behavior for the low-band MN at high-band frequency and vice versa. The two MN branches were then combined together. In [24], again a two CE configuration (called ground plane boosters) was proposed and this time, the two MN branches were connected together

after using notch-filters at each branch (Figure 15). In [25], two CEs which are separately fed have been used for dual-band operation for covering only GSM/DCS/PCS bands. A tunable MN topology for a CE has been proposed in [26] to cover DVB-H, GSM and UMTS frequency bands. In [27-28], two designs have been proposed to additionally cover the low LTE frequency band starting from 700MHz. Antenna proposed in [27] can cover low LTE and UMTS bands (700-960MHz and 1.7-2.2GHz) using a single CE. The matching network at the feed consists of a two-branched MN for dual-band operation, having 9 SMD components. In [28], the coverage band is increased as 700-960MHz and 1.7-2.7GHz, covering whole 4G communication band. The antenna is like a dual-feed structure, where the low-band (LB) feed uses the radiating element as a CE, whereas the high-band (HB) feed has an offset in the excitation region, using also the resonant properties of the radiating element (Figure 16). These two feeds are combined together to obtain a single feed, after introducing a two branch MN (of 8 lumped components) where each branch behaves as an open circuit to the other.

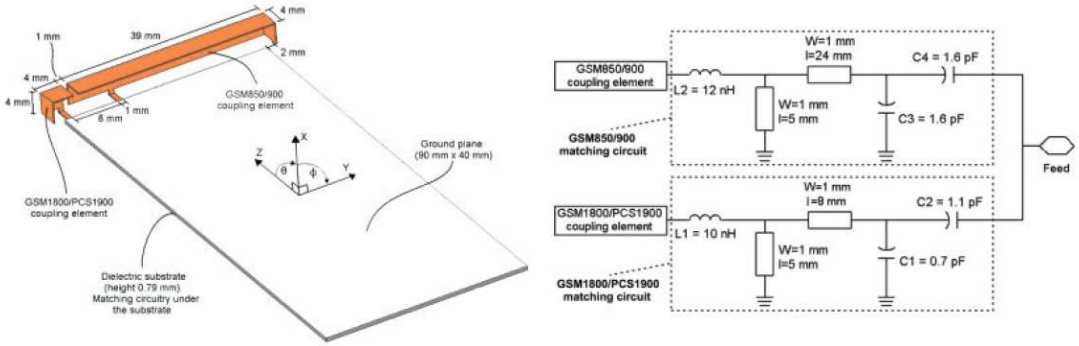

Figure 14 Quad-Band Coupling Element Antenna Structure in [23]

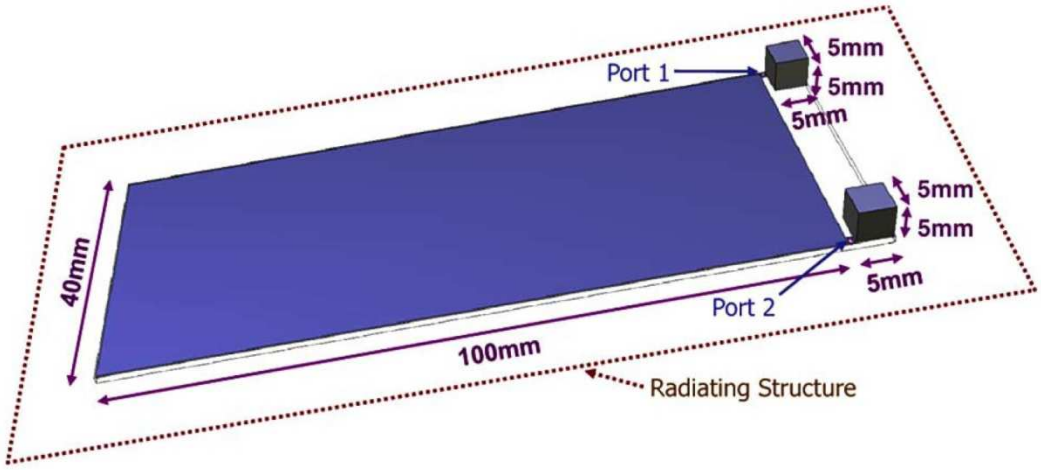

Figure 15 Ground Plane Booster Antenna in [24]

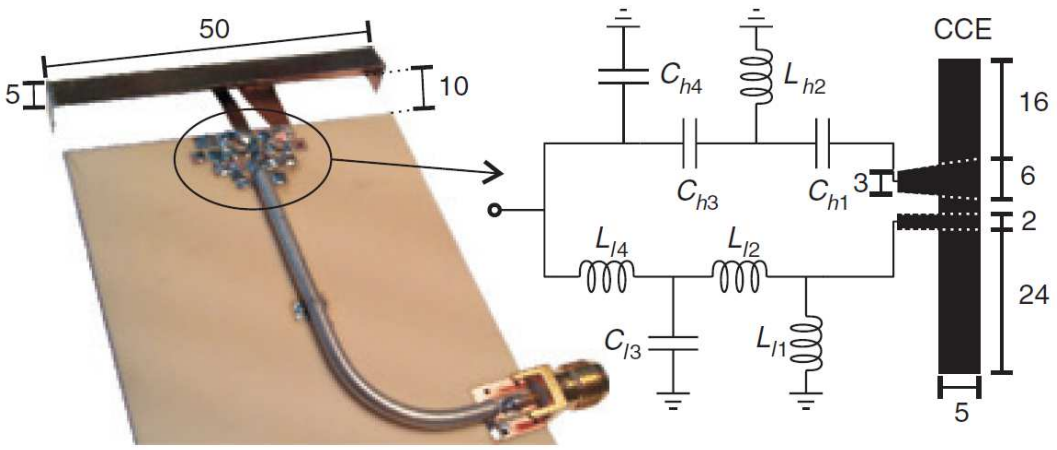

Figure 16 Multi-Band Mobile Terminal Antenna in [28]

7. ANTENNAS USING PARASITIC ELEMENTS FOR BANDWIDTH ENLARGEMENT

For operation in the main 4G frequency bands, wideband antenna behavior is needed covering 30% bandwidth in the low-band (700-960MHz) and 45% bandwidth in the high-band (1.7-2.7GHz). The coverage of the low-band (LB) is more challenging since the antenna is electrically small and thus the bandwidth potential is limited. For bandwidth enhancement, antennas using different kinds of parasitic elements have also been proposed in literature [29-38]. In these antennas, the main methodology is to capacitively excite a longer metal strip connected to the ground plane through a shorter driven strip. In [29-30], a monopole driven strip is used to excite a longer parasitic element connected to the ground plane on one edge. The LB coverage is mainly achieved by the capacitive excitation of the parasitic element and the HB is covered using the driven monopole and the higher order resonances of the parasitic strip. In [31-32], the parasitic element was connected to the ground plane on both edges forming a loop, again excited by a driven strip on the other side of the PCB (Figure 17). To increase the length of the loop parasitic, meandering was used making the design more complex and sensitive to manufacturing tolerances. In the antennas proposed in [33-36], casings for the PCB were also taken into consideration, whose loss tangent vary between 0.01-0.02 (Figure 18). In this way the advantage of dielectric loading was used to be able to achieve LB coverage more comfortably. Dielectric casings were also used in the antennas proposed in [37-38], with loss tangents between 0.02 and 0.06. Although they are called as PIFAs with coupled feeds, the main idea is the same as using a monopole like exciter (without ground connection), to excite capacitively a metal strip connected to ground in one end. The possible drawback of these antennas is the relatively complicated design and complicated antenna layout, which may be a problem when the manufacturing errors are concerned.

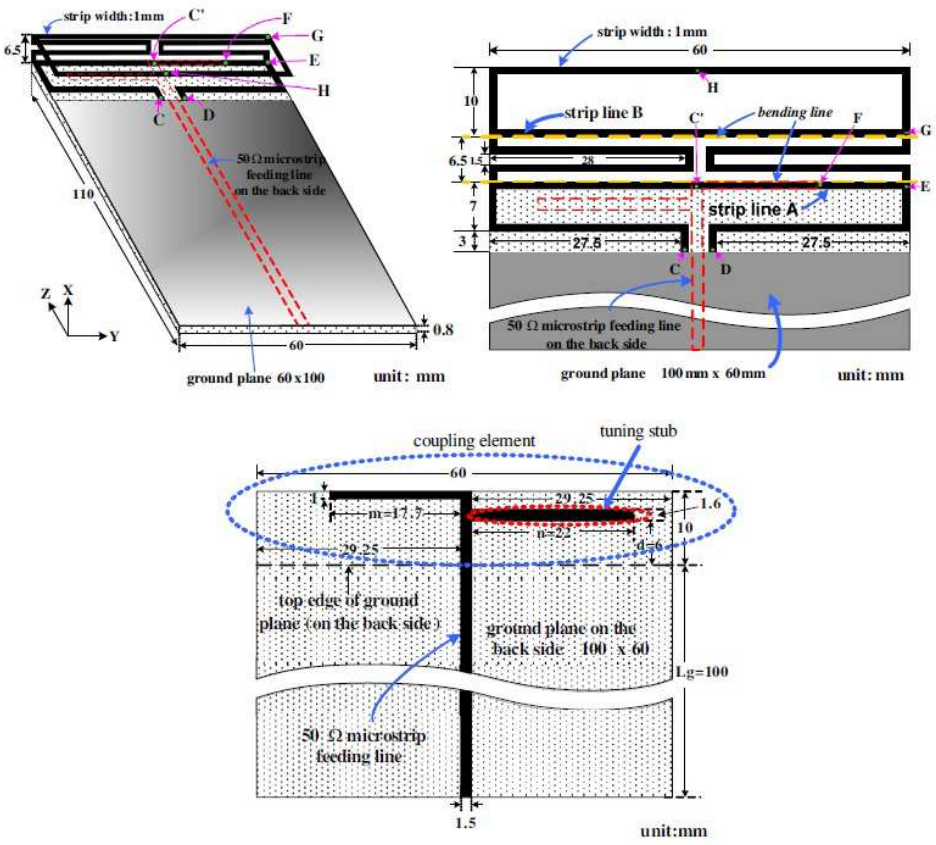

Figure 17 Meandered Loop Antenna in [32]

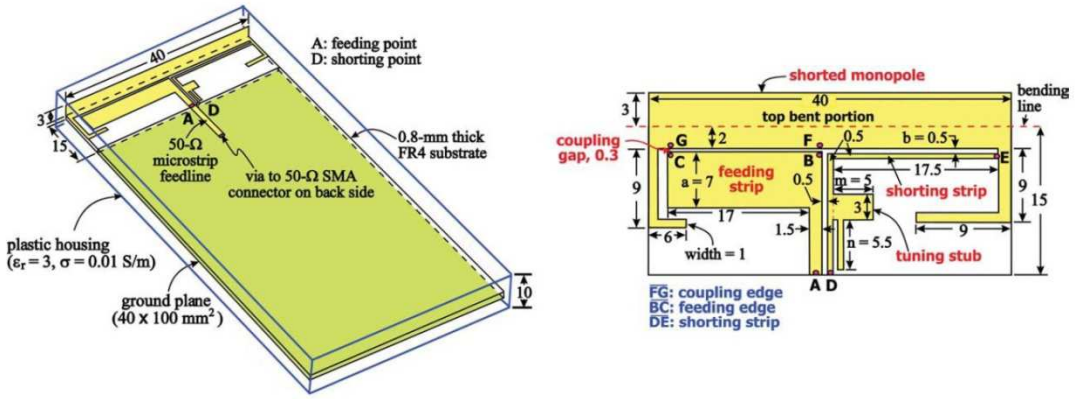

Figure 18 Small Size 11 Band Antenna in [34]

8. SINGLE-FEED AND MULTI-FEED ANTENNA STRUCTURES

The majority of the commercial mobile phones in the market use a single feed antenna design. A study about the advantages and disadvantages of using multi-feed antenna designs compared to single feed antennas was done in [39]. In the current single feed designs, an antenna which is able to cover the required standard frequency ranges (for example GSM850/900, DCS/PCS) is connected to a "Single Pole Multi Throw" antenna switch in the RF front end. According to the communication standard used at the instant, the output of the switch (considering the receive scenario) is changed to connect the signal from the antenna to the appropriate duplexer.

With the introduction of LTE-A, it is possible to receive cellular communication data over two different frequency bands (carrier aggregation), meaning there can be more than one receive chain in the front-end simultaneously. For the inter-band carrier aggregation, the signals picked up by a single feed antenna (for example one at 940MHz and one at 2150MHz) should be transmitted to the front end at the same time, which is not possible with a "Single Throw" switch. The switch should have a logic to have two active throws at the same time, which may results in worse linearity performance and higher insertion loss. In addition to these, special care should be given to prevent the two different front-ends (duplexers) from loading each other since they will be connected together at the switch.

Another solution different from having more than one active throws at the antenna switch, is to add a diplexer just after the antenna in the receive path, that will direct the different frequency signals to the appropriate duplexers, for inter-band carrier aggregation. However this will also result in increased insertion loss in the front-end.

Al alternative solution to overcome these complexities using single-feed antennas for inter-band carrier aggregation is to use multi-feed antennas, for example one covering the lower frequency band (700-960MHz) and other for higher frequency band (1.7-2.7GHz). In this way, the necessity of using a complex antenna switch or a diplexer is overcome for inter-band carrier aggregation. The drawback of this approach is the isolation between the antenna feeds. Isolation should be as high as possible (targeted better than 10dB as a rule of thumb), since it can otherwise cause efficiency losses due to unwanted coupling from one antenna to the other. Obtaining good isolation is also challenging, since the space reserved for the antenna is small in terms of wavelength (especially in lower frequencies) and the separate antennas have to be placed physically close to each other.

9. MOTIVATION OF THE THESIS

This study was done within the scope of the ARTEMOS project, a European Union project with a duration of 36 months. The main aim of the project is to obtain frequency agile, multimode, multi standard RF architectures for mobile communication devices, to decrease the bill of materials and cost of the systems. This target is represented in Figure 19, where the number of RF front-ends for several different communication standards in a typical system nowadays is reduced by proposing tunable front-end modules. Another aim of the project is to design and manufacture tunable components, which are BST capacitors, MEMS switches and MEMS varactors, to be delivered to the partners working on the RF front end and the antenna element.

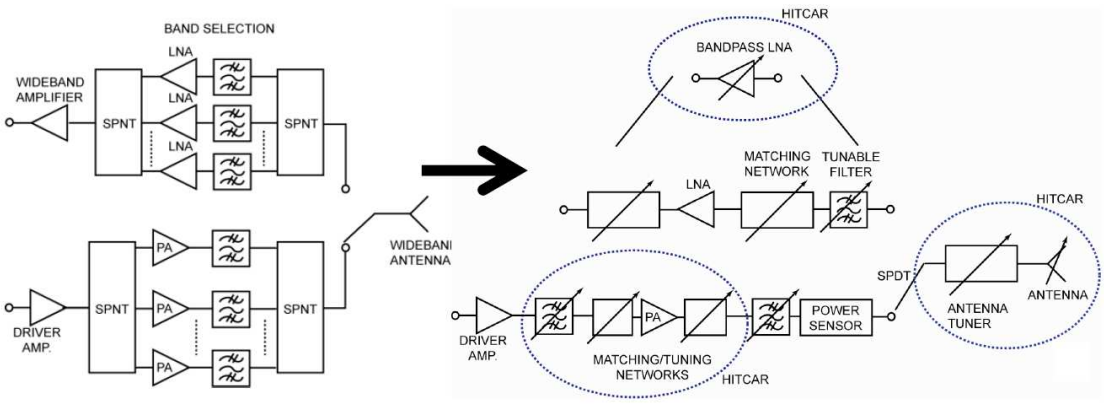

Figure 19 Tunable RF Front End Targeted in ARTEMOS Project

Keeping this in mind, the aim of this study is to design, manufacture and test some novel antenna topologies for handheld mobile terminals. Particular effort is to be focused on either covering the target bands with very simple antenna layouts or achieving lower reflection coefficient in the target band compared to the antennas in the literature mentioned before. Another aim is to investigate the effect of the user (hand and head) on the antenna performances and propose a novel antenna topology which minimizes these effects. Both passive and active antenna topologies are to be evaluated for the coverage of 4G frequency bands, decided as:

-Low-Band (LB): 700-960MHz
-High-Band (HB):1700-2700MHz.

With this target, the frequency bands of the communication standards low-LTE, GSM850/900, DCS/PCS, WLAN2400 and high-LTE will be covered. As a rule of thumb, -6dB in free space is accepted as the target reflection coefficient for the antennas. For the maximum limit of the antenna height, 5mm is taken, to be in compliance with the trends of current smart phone designs. For the same reason, maximum dimensions around 120mm X 60mm (for length and width respectively) was accepted for the PCB that will support the antenna. For the feeding scheme, both single-feed and multi-feed antennas were evaluated.

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CHAPTER II
TUNABLE ANTENNAS

This section presents several tunable antenna topologies, designed for operation in the 4G-LTE frequency bands (i.e 700-960MHz, 1.7-2.7GHz). The antennas are designed and manufactured using tunable components provided by the ARTEMOS partners. To obtain tunability in the matching response, MEMS switches and BST capacitors have been considered.

1. ANTENNA DESIGNS USING MEMS SWITCHES

The utilization of MEMS switches in reconfigurable antenna design has been a popular topic in recent years, especially due to their low loss in the microwave frequency bands. The reconfigured antenna properties of the these antennas have a diversity, like pattern reconfiguration (both in single and array antennas), switching between two operating frequency bands, tuning the matching level by MEMS capacitors (varactors) and changing the properties of antenna feeding networks [1]. For example in [2], MEMS switches were used to change the ground connection point of a PIFA to achieve polarization reconfigurability. In [3-4], the switches were used to obtain reconfigurability in the operating band. Figure 20 shows the proposed reconfigurable MN topology in [3], together with the manufactured prototype in the same study.

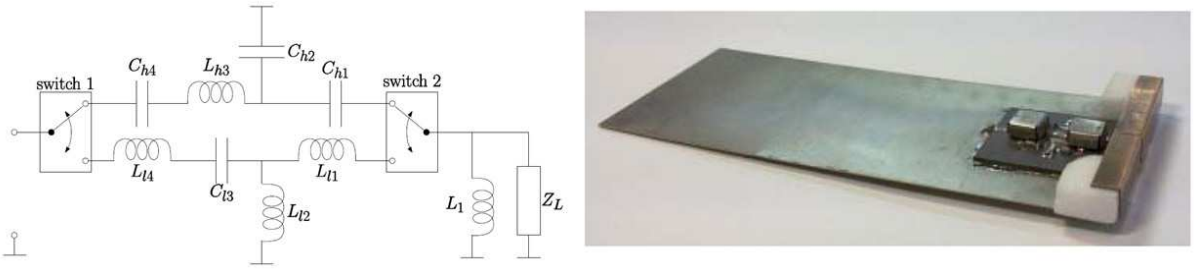

Figure 20 MN Topology and Manufactured Prototype in [3]

There are a number of COTS products available on the market, some of which are given in [5-6] and whose RF performance specifications are mentioned in Table 3.

Table 3 Specifications of Some Available MEMS Products

	MEMS-1	MEMS-2
Frequency Bandwidth	DC-40GHz	DC-10GHz
Insertion Loss at 2GHz	-0.29dB	-0.25dB
Isolation at 2GHz	-34dB	-45dB
Maximum RF Input Power	40dBm (cold sw.)	36dBm
Input IP3	>65dBm	
Switching ON Time	<10 µs	<100 µs
Control Voltage	90V	35V

In this study, the MEMS switches have been evaluated to achieve band-switching between the Low-Band (700-960MHz) and High-Band (1.7-2.7GHz) of 4G communication standards. Two antenna topologies are proposed, each using a parasitic strip connected to ground for the enlargement of matching bandwidth. The two topologies are different in the way the MEMS switches are exploited. In the first design, two MEMS switches are used to select between one of the two matching network

(MN) branches, like in [3] and [4]. The second design uses only one MEMS switch, to select between one of the driven strips.

The MEMS switches used in this study are designed and manufactured by DELFMEMS [7], who is also a partner in the Artemos project. DelfMEMS technology is different from the other RF MEMS switches on the market, i.e. classical cantilevers (clamped-free membrane) or bridges (clamped-clamped membrane). Indeed, based on a state-of-the-art IP, DelfMEMS switches enable high performances with their specific mechanical structures featuring two main innovations: a movable membrane without anchors in the substrate and two forced on/off states to obtain a push-pull behavior. The membrane used for the electromechanical function is simply supported by two pillars, and two sets of electrodes placed either side of the pillars allow the actuation of the device. As illustrated in Figure 21(on left), the two forced states are obtained when the electrostatic forces are generated with the different electrodes, respectively internal/external electrodes for a ON/OFF state. Such operation ensures good contact forces in the ON state thus reducing the contact resistance, and provides high restoring forces for the OFF state to counter potential stiction or self actuation phenomena while ensuring important isolation.

In addition to good RF characteristics, this new mechanical structure demonstrates high reliability performances mainly because of its anti-stiction capabilities and the limitation of bounces thanks to a smart use of the different internal/external electrodes.

The insertion loss of the devices is measured to be between -0.2 and -0.25dB at 2GHz (Figure 21, on right). A high level of isolation is observed in OFF state, better than 45dB below 3GHz thanks to isolated RF and DC paths and because a gap larger than the thickness of the sacrificial layer is obtained between the contacts placed on the membrane and those on the transmission line

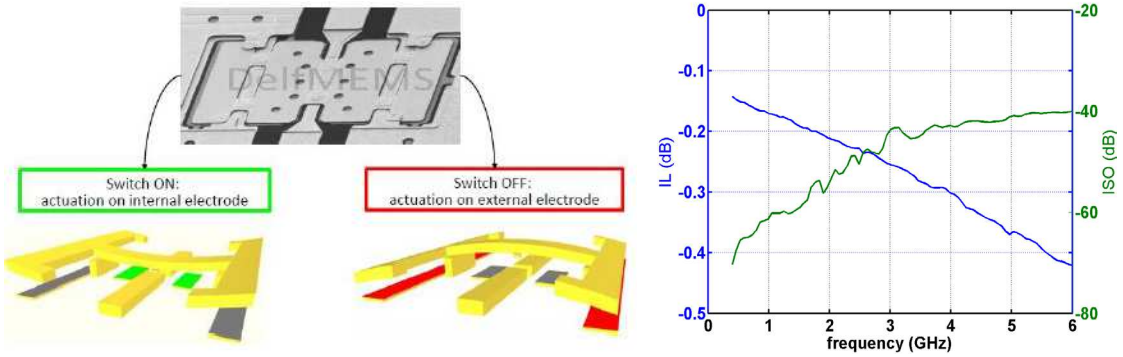

Figure 21 ON/OFF Switching Topology and Performance of MEMS

Table 4 Performance Summary of MEMS switches from DELFMEMS

Parameters	Value
Frequency bandwidth	0.3-5 GHz
Insertion loss	0.2-0.25 dB @2GHz
Reflection coeff.	< -15 dB @2GHz
Isolation	> 50 dB @2GHz
RF power	+33 dBm for cold switching +20 dBm for hot switching
Input IP3	> 70 dBm
Switching time	<10 µs for actuation voltage greater than 50V
Control signal level	50-70 V
Switching frequency	up to 10 kHz
Electrostatic self-actuation (OFF state)	< 50 V peak peak
Size	3,6 mm²
Packaging type	silicon cap

1.1. ANTENNA DESIGN-I

The proposed antenna layout can be observed in Figure 22. The antenna is directly printed on FR4 substrate, with the PCB dimensions of 120mm X 60mm X 0.8mm. There is a ground plane of length 107mm under the substrate on one side of the PCB, leaving a ground clearance of 13mm for the antenna. The antenna is basically composed of a driven strip, which capacitively excites a parasitic strip connected to the ground plane. The parasitic strip is connected to the ground plane over an inductor (12nH), which increases the electrical length of the parasitic strip. A MN, consisting of two branches is introduced at the antenna feed (Figure 23), to tune the antenna either in the LB or in the HB. According to the states of the two SPDT (Single Pole Double Throw) switches, one of the branches is connected to the antenna feed, which matches the impedance in different frequency bands. The LB MN branch consists of two inductors, whereas the HB MN is a single capacitor.

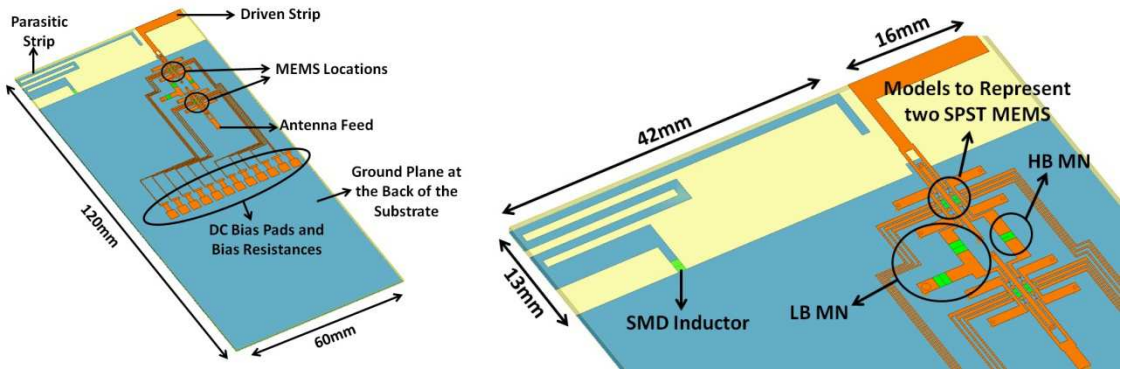

Figure 22 Layout for Antenna Design-I Using MEMS

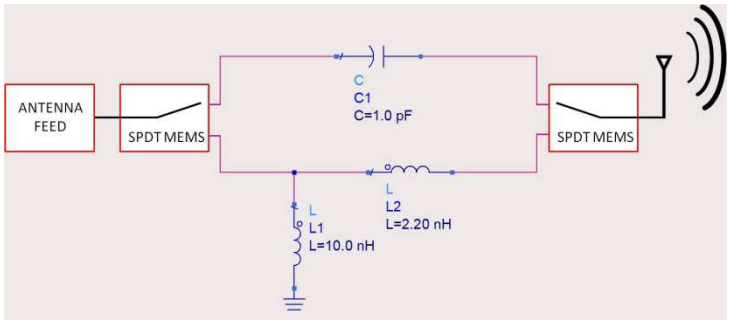

Figure 23 Switchable Matching Network Topology

1.1.1. Effect of the Parasitic Strip over the Bandwidth Potential

The effect of the parasitic strip has been studied in the absence of any matching network at the antenna feed. The simulated antenna input impedance with and without the parasitic strip can be observed in Figure 24, plotted between 700-960MHz. A significant increase can be noticed in the real part of the impedance, with the resonance appearing around 730MHz. Here, the term "resonance" is used to indicate a significant increase in the real part of the antenna (the loop forming in the smith chart), since under normal conditions the real part of the impedance of an electrically small antenna is low. The increase in the real part of the input impedance can be seen in Figure 25, for "with Parasitic" case.

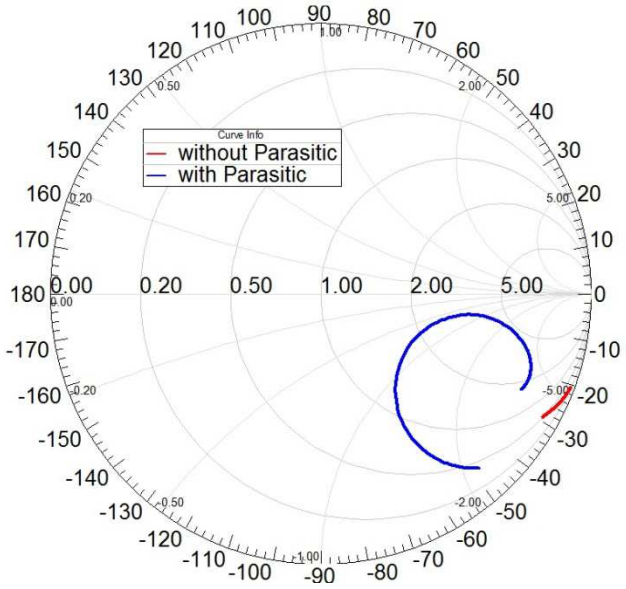

Figure 24 Input Impedance of the Antenna With/Without Parasitic Strip from 700 960 MHz

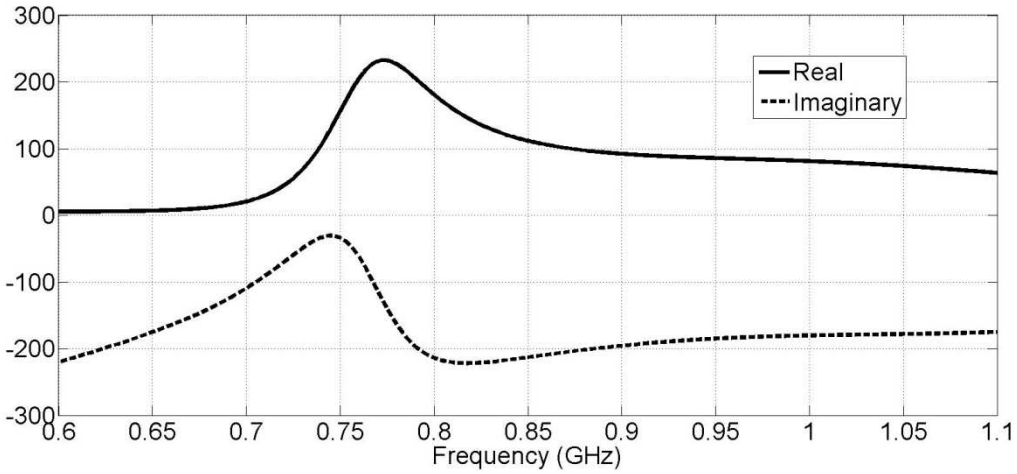

Figure 25 Input Impedance (Real/Imaginary) of the Antenna With Parasitic Strip

The surface current density of the antenna presented in Figure 26 also verifies this statement. At 700MHz, the currents induced on the parasitic strip are strong, creating a current path from the driven strip to the parasitic strip and returning to the feed through ground connection. At 950MHz, the currents are significantly weaker on the parasitic element. As previously discussed, the excitation of the ground plane plays a major role in the whole radiation mechanism of the structure and therefore the radiation efficiency and also the matching coverage of the LB. When the current distribution on the ground plane is investigated, it may be concluded that the first chassis wavemode of the ground plane is excited uniformly at 950MHz.

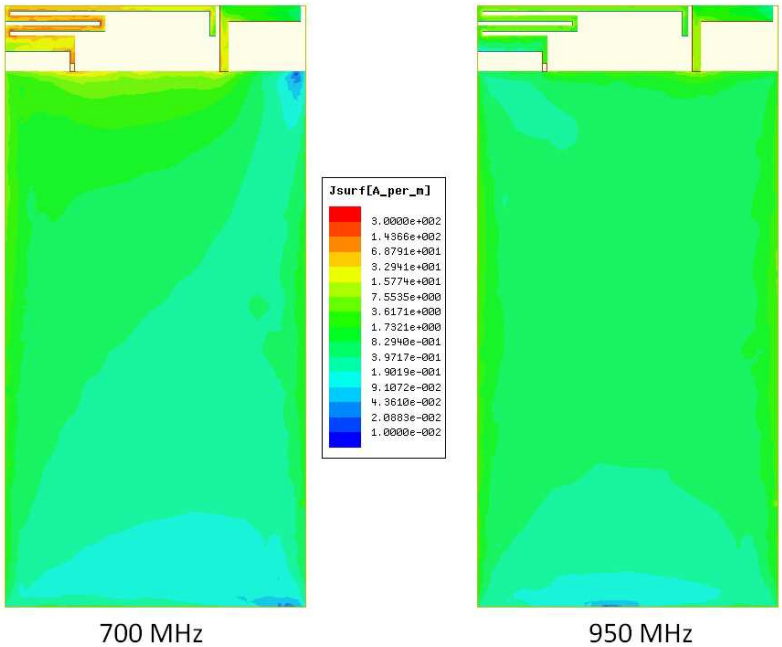

Figure 26 Surface Currents of the Antenna and The Chassis in LB

The bandwidth potential of the two cases (with/without parasitic element) has been investigated using the software Optenni Lab [8]. For this computation, the software optimizes a matching network (composed of two elements) for each frequency point and determines the frequency interval that rests under a pre-defined reflection coefficient value (-6dB chosen here). In this way, a realistic

bandwidth potential value is given instead of well-known, too optimistic theoretical maximum value. It can be seen in Figure 27 that introducing the parasitic strip increases the bandwidth potential considerably in the LB. The computed bandwidth potential gets closer to the target curve (in red), which states the bandwidth potential needed at each frequency point to cover 700-960MHz.The parasitic strip enables a bandwidth potential of approximately 250MHz at the center frequency (830MHz), with a two element MN.

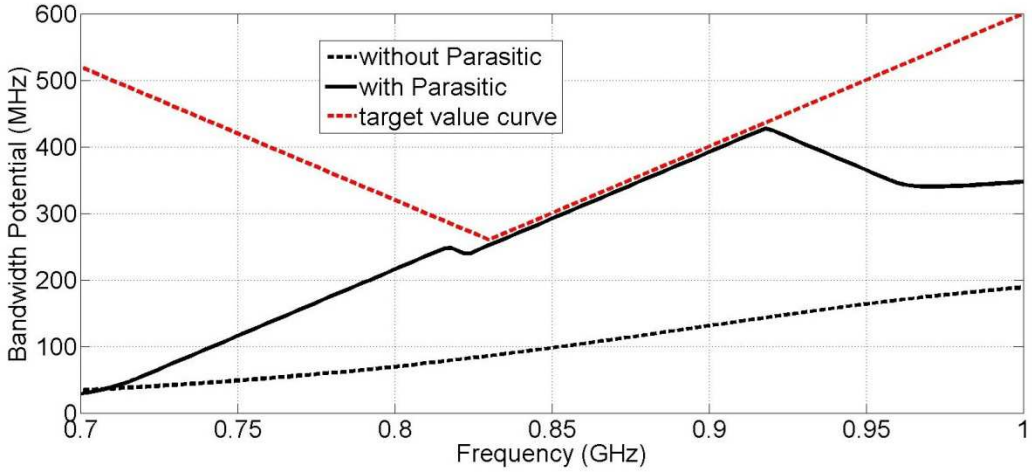

Figure 27 Comparison of BW Potential of the antenna With/Without Parasitic Element

The effect of the parasitic strip for creating an additional resonance in the LB was also investigated through circuit models. Circuit models were proposed for the driven strip in the absence of parasitic strip (Figure 28 on left) and for the parasitic strip alone when it is directly fed (Figure 28 on right). The circuit models for these two configurations are modeled as a series RLC circuit with a shunt capacitor in front which represents the capacitive coupling to the ground plane. A good correspondence between the simulated EM model and the circuit models can be observed (all smith chart curves are plotted between 700-960MHz).

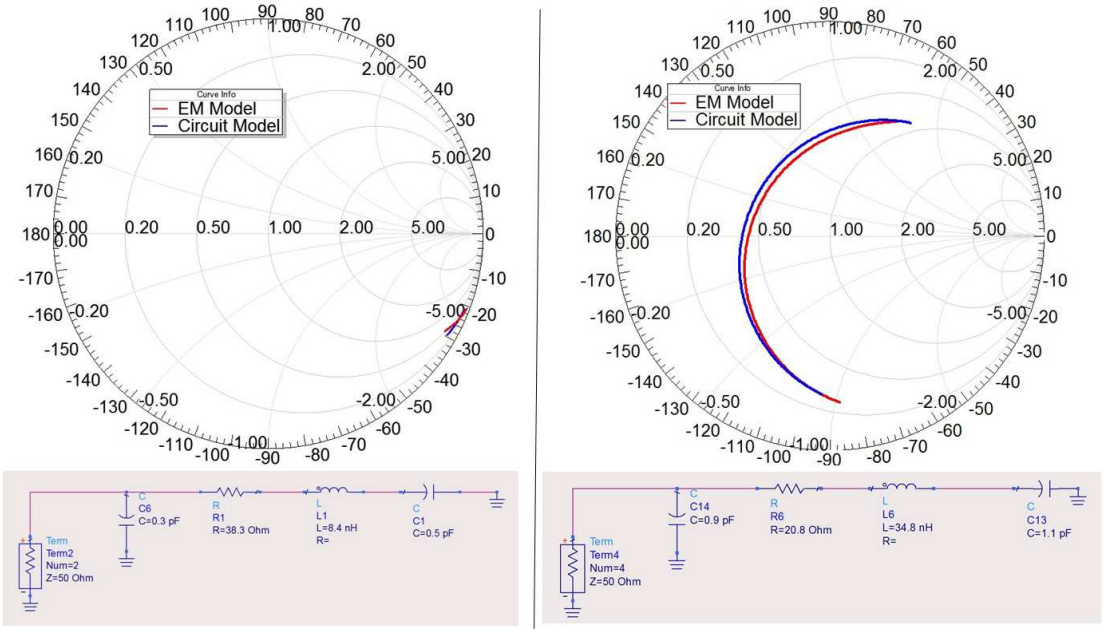

Figure 28 Circuit Model of the Driven (on left) and Parasitic Strips (on right)

Using these two circuit models as building blocks, the circuit model in Figure 29 is obtained. The circuits for the driven strip and parasitic strip are marked with the black and red circles respectively. A transformer model is used to stand for the capacitive coupling between the strips. The creation of the resonance in the LB can be observed for the circuit model, making a good compliance with the EM simulations.

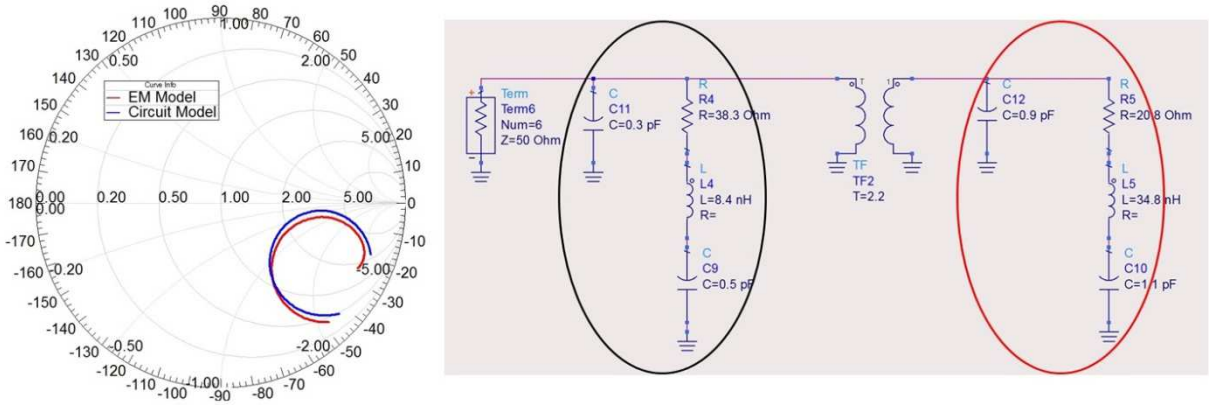

Figure 29 Circuit Model of the Combination of Driven and Parasitic Strips

The effect of the parasitic element in the HB (1.7-2.7GHz) can be observed in Figure 30. With the introduction of the parasitic element into the "driven strip only" case (red curve) two additional resonances are created around 1.85GHz and 2.4GHz. If the current distributions are observed at these two frequencies, it can be seen that the parasitic element is excited strongly in two different operating modes. Considering this fact, it can be concluded that the role of the parasitic strip is not only in the LB. It also contributes to the current distribution in the HB frequencies.

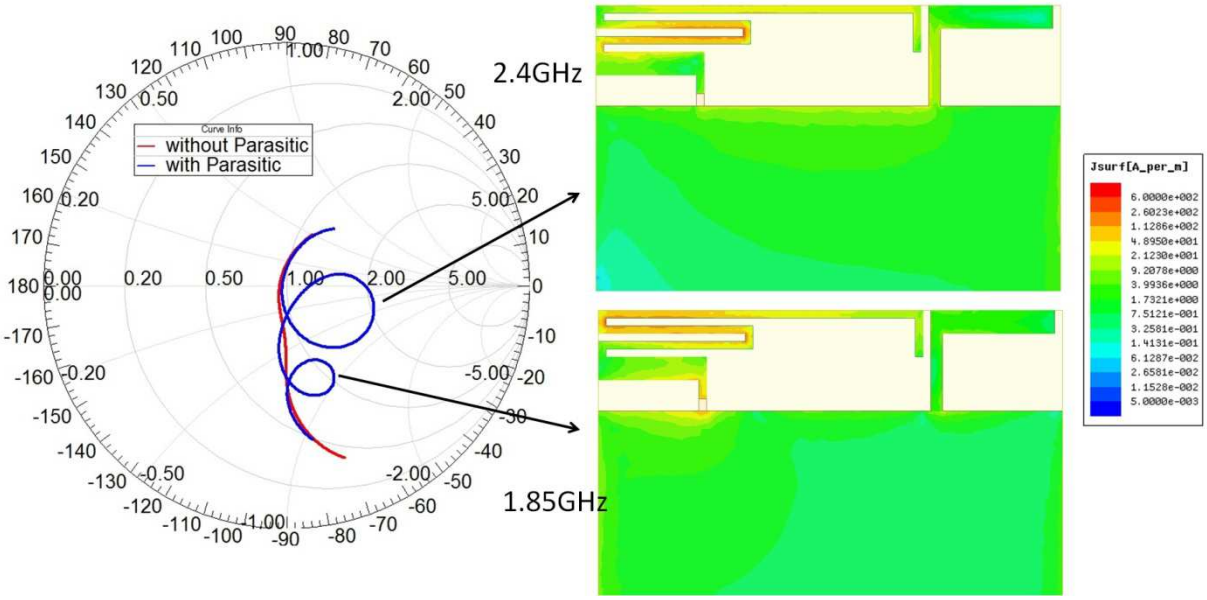

Figure 30 The Effect of the Parasitic Strip in the HB

1.1.2. Antenna Simulations With MEMS Switches and Matching Network

The antenna whose layout has been given in Figure 22 was simulated using Ansoft HFSS [9]. In the full EM simulations, the MEMS switches and SMD components were replaced by simulation ports and measured S-parameters from the manufacturers were used at those ports with a circuit post-processing simulation done in Agilent ADS [10], to obtain more realistic results. The DC bias pads for the MEMS switches and the wirebondings (to connect the MEMS die to the traces on the PCB) were also taken into account in the full EM simulation(Figure 31).

The main criteria when optimizing the antenna was to first obtain enough bandwidth potential for the LB coverage. The main antenna parameters that were tuned for this, were the coupling gap between the driven strip and the parasitic strip, the length of the parasitic strip (that changes the frequency of the resonance) and the inductance value on the ground connection of the parasitic strip (tuning the electrical length and thus resonance frequency). Since the antenna is longer in terms of wavelength, the coverage is simpler for the HB.

The simulated input impedances of the final optimized layout, with and without MN can be seen in Figure 32, plotted for LB (700-960MHz) on left and plotted for HB (1.7-2.7GHz) on right. It is worthwhile to note that the SMD component values in the MN were tuned taking into account also the transmission lines between SMDs and also the wirebondings. To show the effect of the transmission lines between MN elements, simulation were also done by connecting the lumped elements directly in the circuit simulator (dashed curves in Figure 31) and neglecting the wirebondings. It is seen that the transmission lines might have a strong effect depending on the input impedance of the antenna before the MN.

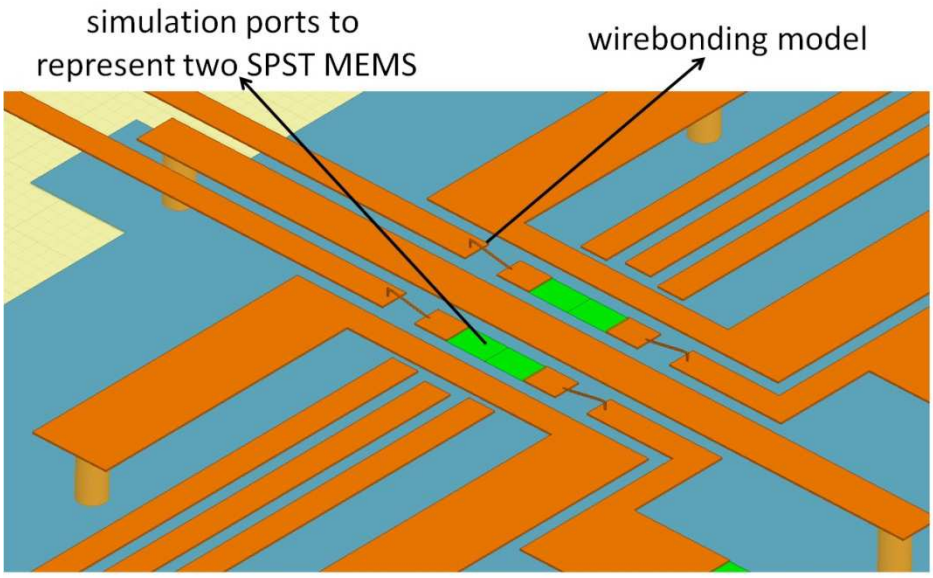

Figure 31 Simulation Model Showing Wirebondings and MEMS Modelling

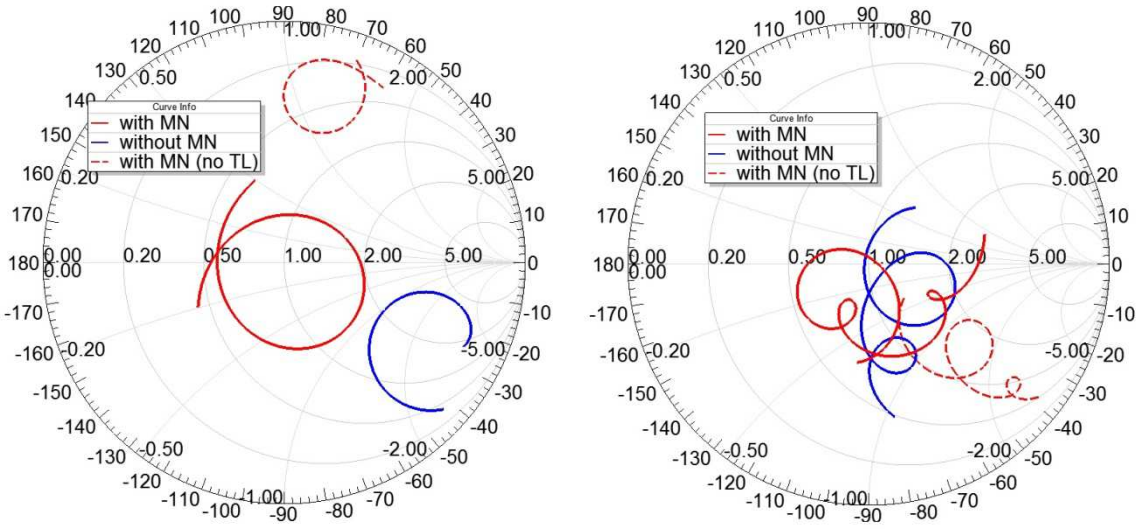

Figure 32 Simulated Antenna Impedance With/Without MN (LB on left, HB on right)

The simulated reflection coefficient of the antenna can be seen in Figure 33. Without any MN, the antenna is far from being matched in the LB (red dotted curve). With proper switching of the MEMS switches to select the LB MN (black curve), the antenna can be matched in the LB with a reflection coefficient below -6dB . Likewise, by choosing the HB MN branch, the antenna has a reflection coefficient below -6dB in the HB.

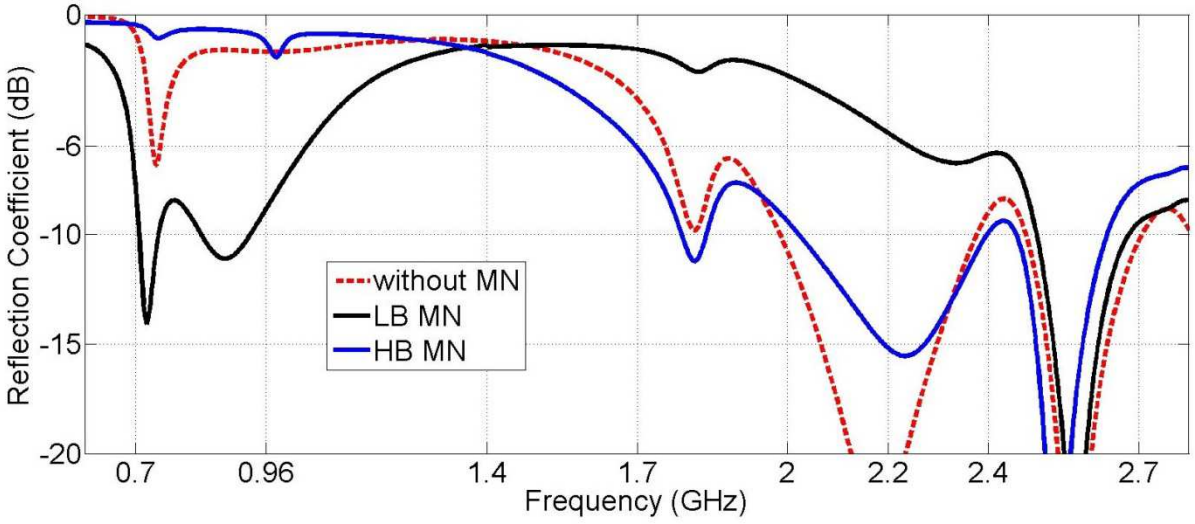

Figure 33 Simulated Reflection Coefficient of the Antenna With/Without MN

1.1.3. Measurement Results for Antenna Design I

The proposed antenna has been manufactured on FR4 substrate and the integration of the MEMS switches to the PCB was done using wirebonding (Figure 34).

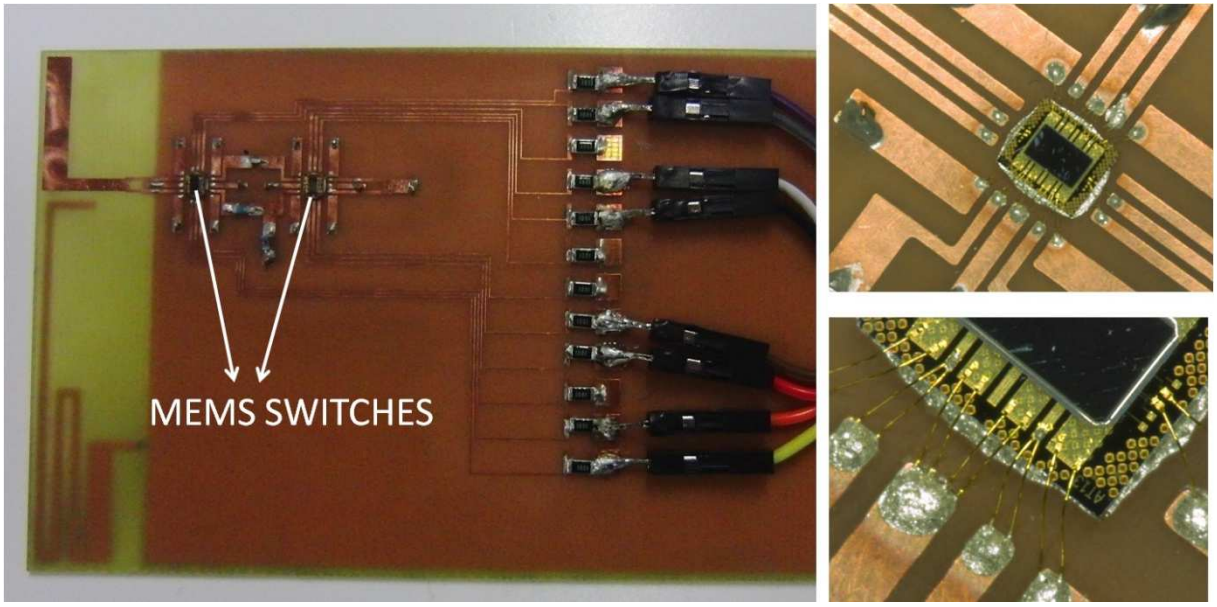

Figure 34 Pictures of the Manufactured Prototype for Antenna Design I

Due to the lack of encapsulation of the MEMS dies, problems were faced during the measurements mainly about the MEMS reliability and repeatability of the measurements. For this reason, as a backup solution, passive prototypes were manufactured where the MEMS switches are by-passed using soldered wires for ON state and left open for OFF state. The comparison of the measured reflection coefficient (for passive prototypes) and the simulation results can be seen in Figure 35 and Figure 36. A good correspondence is observed between simulations and measurements in terms of matching and input impedance behavior through both LB and HB.

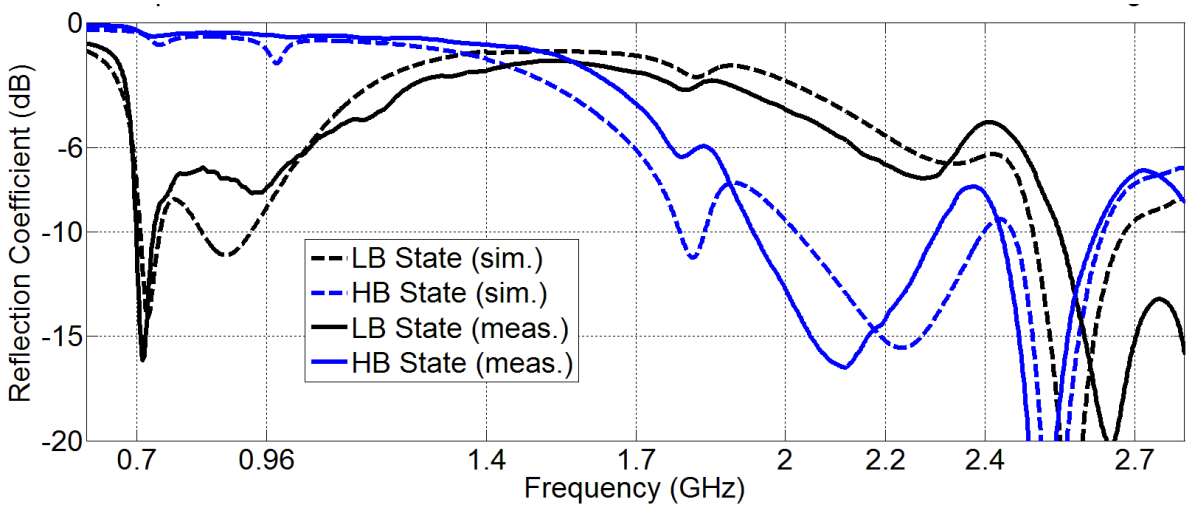

Figure 35 Comparison of Simulated and Measured Reflection Coefficient for Antenna Design I

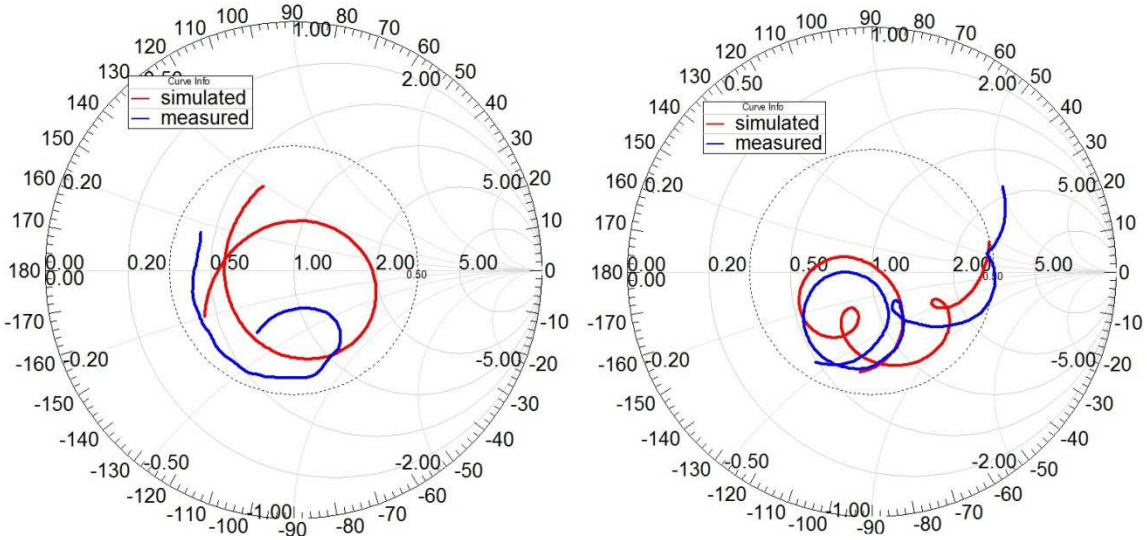

Figure 36 Comparison of Simulated and Measured Input Impedance for Antenna Design I (LB on left, HB on right)

The total efficiency of the passive prototype was measured in a Satimo Starlab station [11], in LB and HB (Figure 37). The measured total efficiency (including mismatch losses) can be seen in Figure 38 to be between -3.5dB and -1.5dB in the LB and between -3.5dB and -1.1dB in the HB.

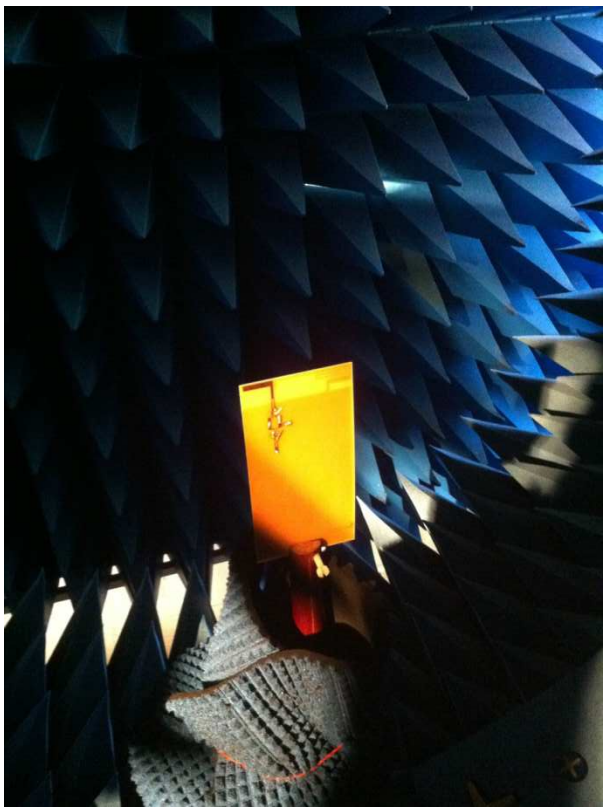

Figure 37 Efficiency Measurement Setup

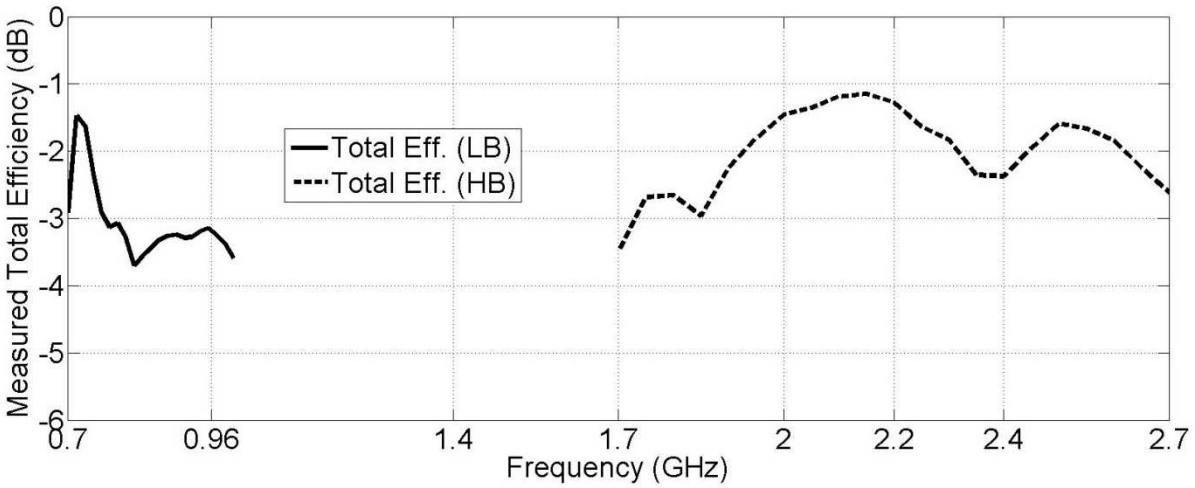

Figure 38 Measured Total Efficiency for Antenna Design I

The comparison of the simulated and measured 3D Gain pattern for the antenna in three different frequencies can be observed in Figure 39. A good compliance is seen between the simulated and measured patterns. In the LB, a dipole type radiation is observed as expected, since the ground plane is the main radiator. The antenna becomes more directive in the HB, as the electrical length of the antenna increases. The vertical polarization (when the PCB is held vertically) is dominant in both bands, where the measured maximum gain differs by more than 10dB in the LB and around 2-4dB in the HB when compared to horizontal polarization.

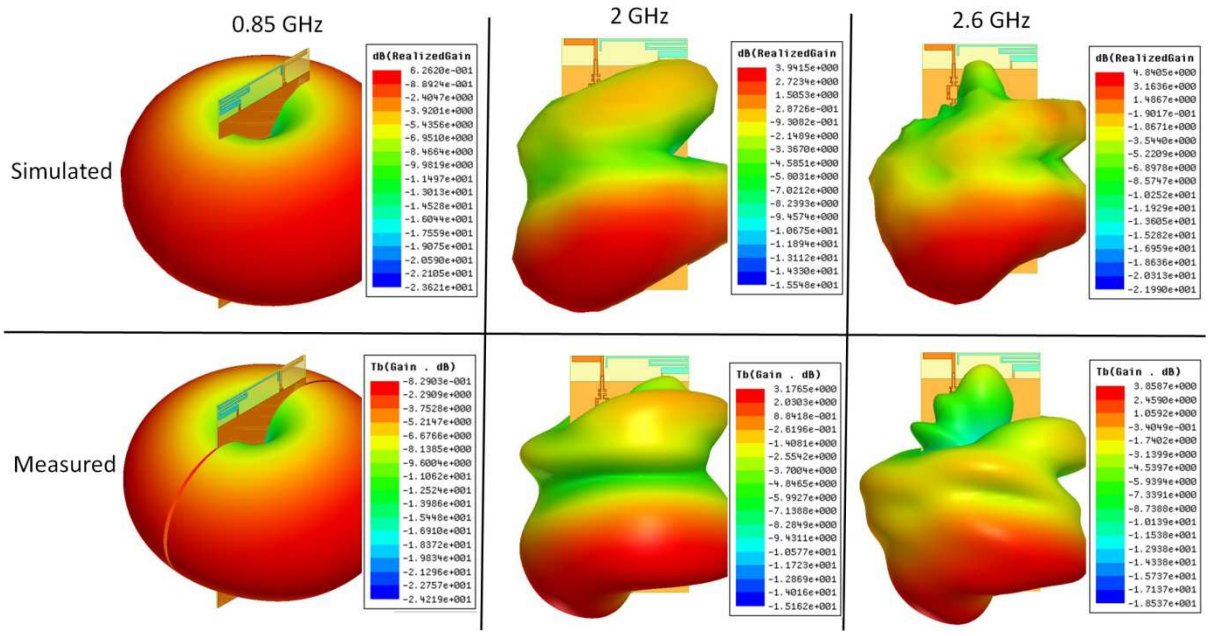

Figure 39 Simulated and Measured 3D Gain Patterns for Antenna Design I

A situation about the radiation efficiency was observed, having a closer look at the HB performance of the antenna. The simulated radiation efficiency is compared with the radiation efficiency calculated from the measured total efficiency (using the measured reflection coefficient) in Figure 40. Firstly, there is a difference of approximately 0.5dB between simulated and measured radiation efficiency, whose possible cause is the accuracy of the measurement set-up given as +/-1dB

in the product datasheet. Besides from that, a slight drop in the radiation efficiency can be seen around 1.85GHz and 2.45GHz both in simulations and measurements. As previously stated in Figure 30, these two frequencies correspond to the strong excitation frequencies of the parasitic strip in the HB. The cause of this radiation efficiency drop was further investigated and will be explained in Section 1.3 of this chapter because it's an important issue in many practical antenna designs for mobile phones.

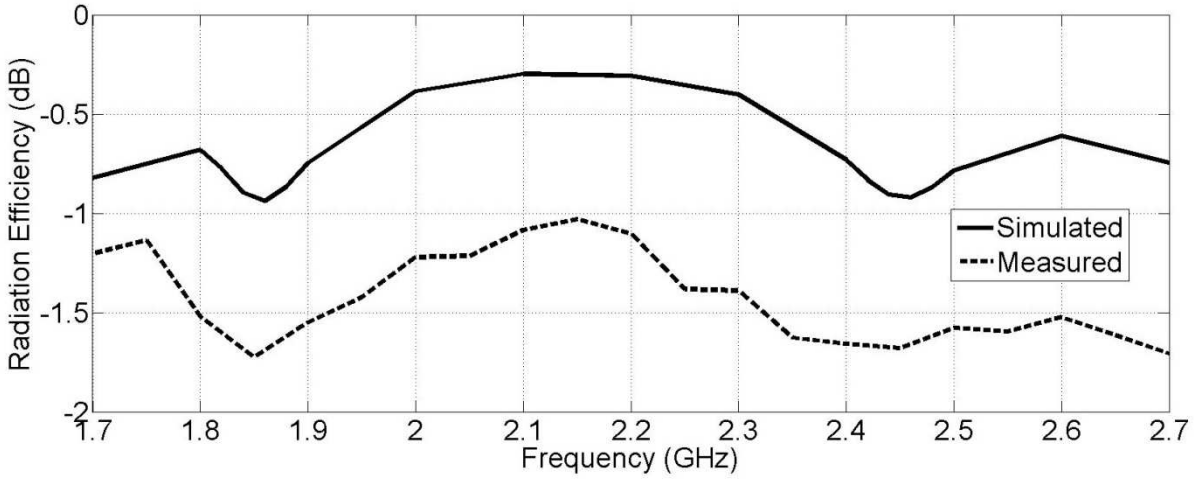

Figure 40 Simulated and Measured Radiation Efficiency in HB for Antenna Design I

1.2. ANTENNA DESIGN-II

A second alternative antenna design was also proposed, using MEMS switches again for band-switching between the LB and HB. The antenna excitation methodology was changed from selecting between two MN branches to selecting between two driven strips. The proposed antenna layout can be seen in Figure 41. The antenna is directly printed on the FR4 substrate having dimensions of 115mm X 60mm X 0.8mm, with a ground clearance of 15mm for the antenna. The antenna consists of two driven strips, one for LB coverage and one for HB coverage. To increase the bandwidth potential in the LB, a parasitic element (printed under the substrate) was again used, which is capacitively excited by the LB driven strip and connected to the ground plane over an SMD inductor (18nH) on the other end. The inductor used in the ground connection helps to increase the electrical length of the parasitic strip, thus decreasing the resonance frequency created by the parasitic strip. The number of SPDT MEMS switches is reduced to one (instead of two in the former case) and the switch is used to direct the excitation current either through the LB path or the HB path as seen in Figure 41. When the LB driven strip is excited, the parasitic strip is also capacitively excited, enabling a higher bandwidth potential in the LB.

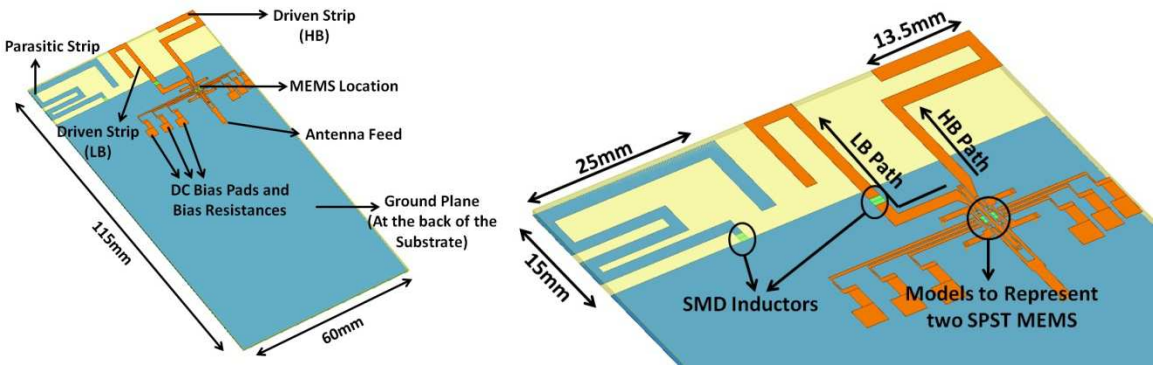

Figure 41 Layout of the Antenna Design-II Using MEMS Switches

1.2.1. Antenna Simulations With MEMS Switches and Matching Network

The simulated input impedance when removing the HB driven strip (considering only the LB driven strip with the parasitic element) can be observed in Figure 42 (Blue curve), on the left smith chart (plotted between 700-960MHz). A resonance is created by the parasitic strip similarly to the one created for Design I, which appears as a new loop in the smith chart. The LB response of the antenna is matched to 50Ω using a series inductor (27nH) on the LB driven strip. The red curve in the same figure presents the input impedance after the HB driven strip and the SMD inductor is introduced to the antenna. If the HB response of the antenna for "HB driven strip only" case is observed on the right smith chart, it can be seen that the parasitic strip creates a resonance in this band also at 1.76GHz.

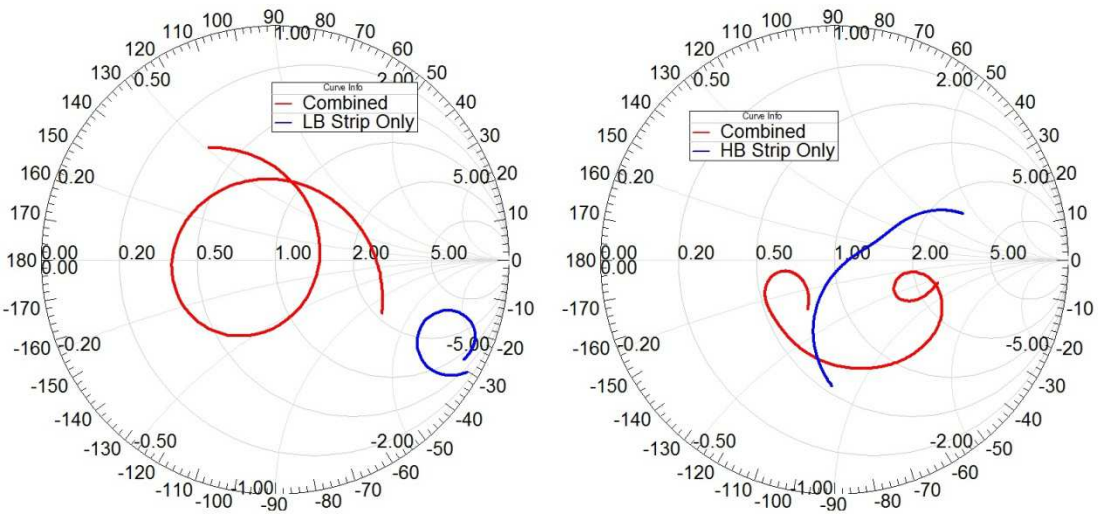

Figure 42 Simulated Input Impedance for Antenna Design II (LB on left, HB on right)

The role of the parasitic strip in the HB can be seen with the surface currents presented in Figure 43. At 1.76GHz, strong currents are excited on the parasitic strip, which appear weaker at any other frequencies (2.2GHz presented here).

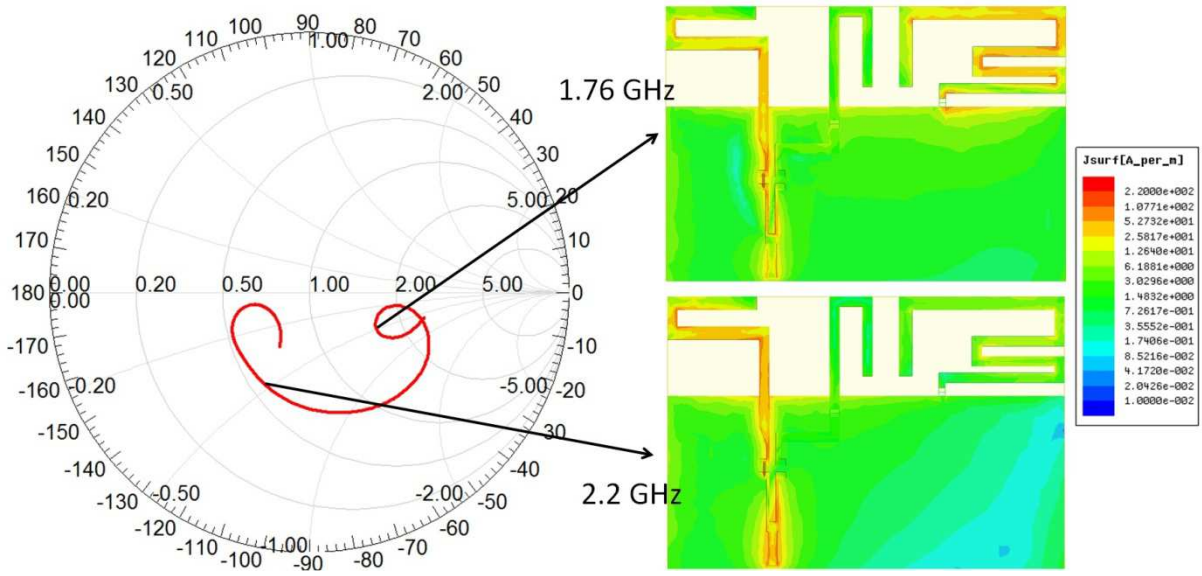

Figure 43 Surface Current Distribution for Antenna Design II in HB

1.2.2. Measurement Results for Antenna Design II

The S-parameter measurements of this prototype have been done both with real MEMS devices and with passive prototypes. In passive measurements, to represent the ON-state of the MEMS, a simple wire was soldered bypassing the MEMS and the connection was left open to represent the OFF-state. The measured reflection coefficient for the antenna in the LB and HB states can be seen in Figure 44. The target bands are covered with a reflection coefficient below -6dB in both passive measurements and real MEMS measurements, using the band-switching topology.

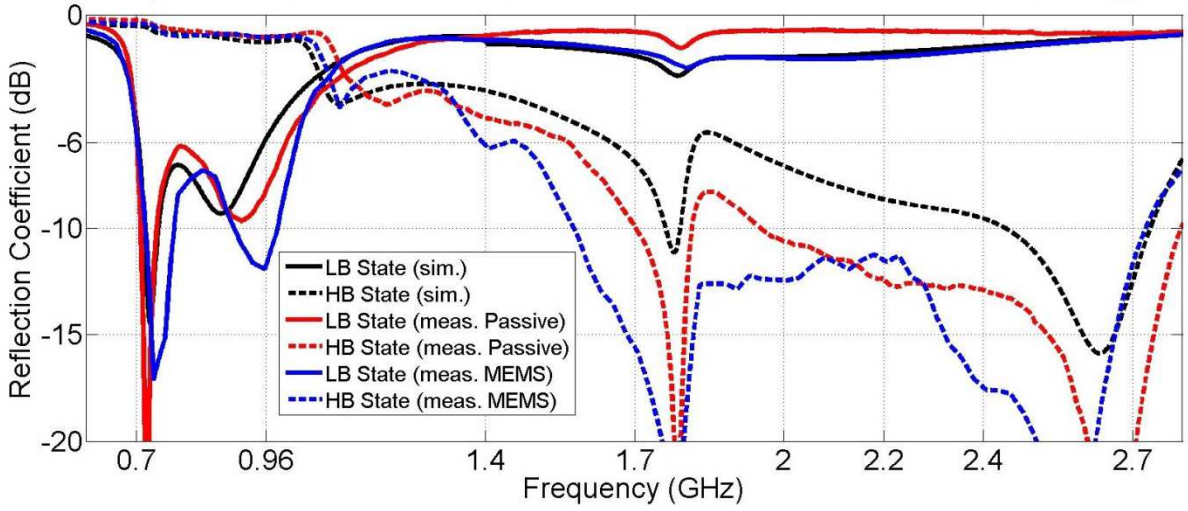

Figure 44 Comparison of Simulated and Measured Reflection Coefficient for Antenna Design II

The simulated and measured (with passive prototypes) input impedance makes a good compliance as seen in Figure 45. Comparing the simulated impedance response with the one measured with MEMS, it can be told that the MEMS switches turned out to be more lossy in the measurements than expected, since the locus in the smith chart is closer to the center (50Ω).

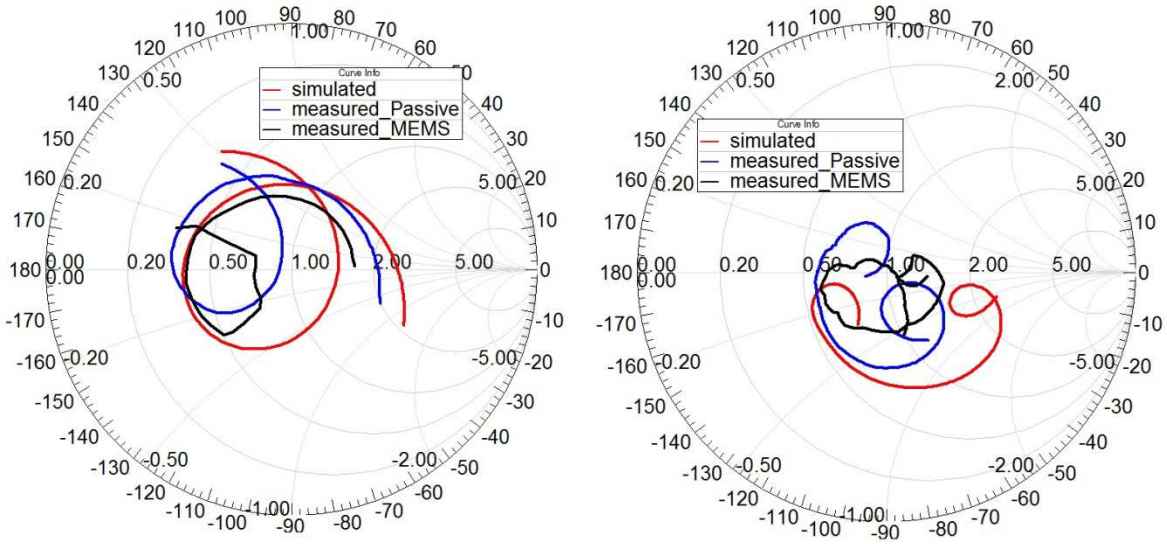

Figure 45 Comparison of Simulated and Measured Impedance for Antenna Design II (LB on left, HB on right)

The total efficiency of the antenna was also measured (for passive prototypes only, due to reliability issues) for LB and HB cases (Figure 47). The manufactured prototype can be observed in Figure 46.There is an efficiency higher than -3dB except around 700MHz in the LB and higher than -2.5dB in the HB.

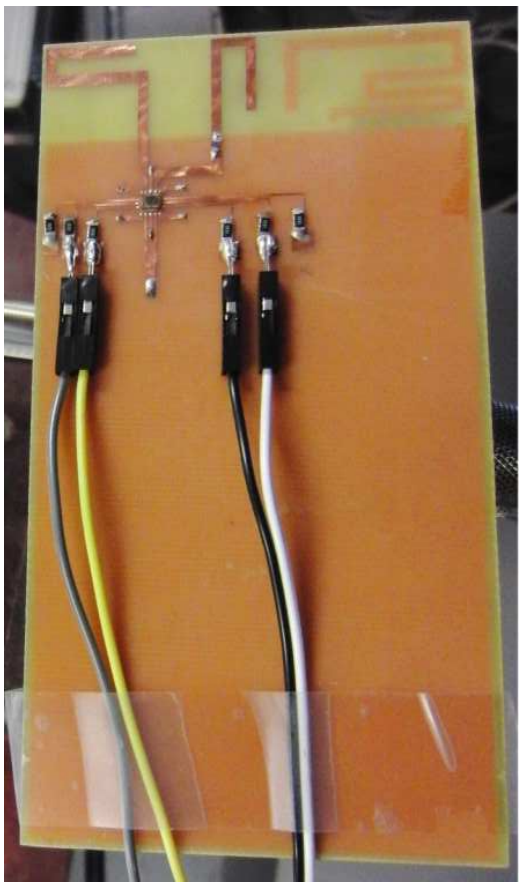

Figure 46 Manufactured Prototype for Antenna Design II

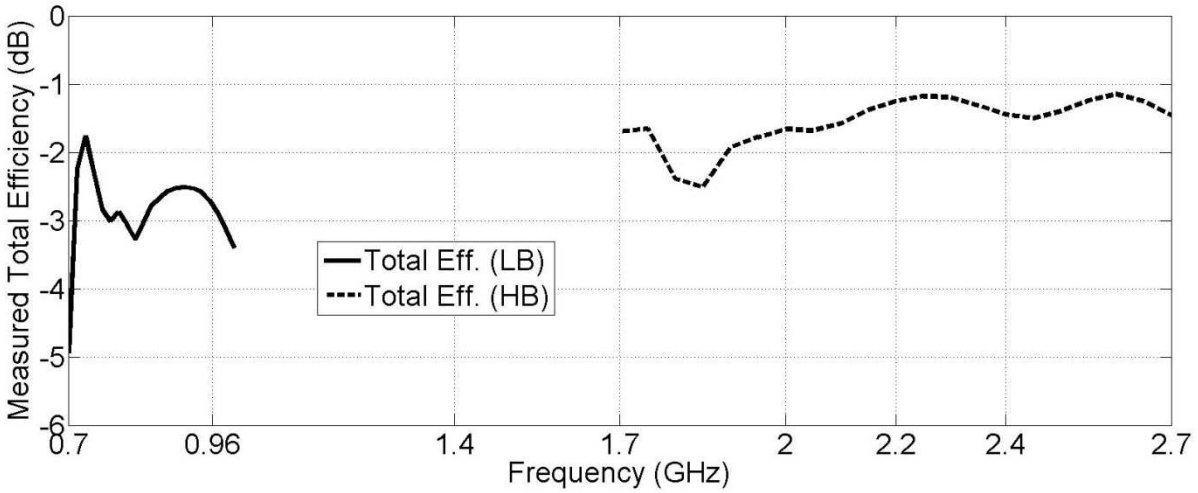

Figure 47 Measured Total Efficiency for Antenna Design II

The measured 3D gain patterns are compared with simulations in Figure 48. A good compliance in terms of pattern shape can be seen between simulations and measurements. As in the previous antenna, the LB pattern is a dipole type pattern and the gain increases in the higher frequency bands. The dominant polarization is again vertical, maximum gain being more than 10dB superior in LB and 1.5-2dB superior in HB when compared to horizontal polarization.

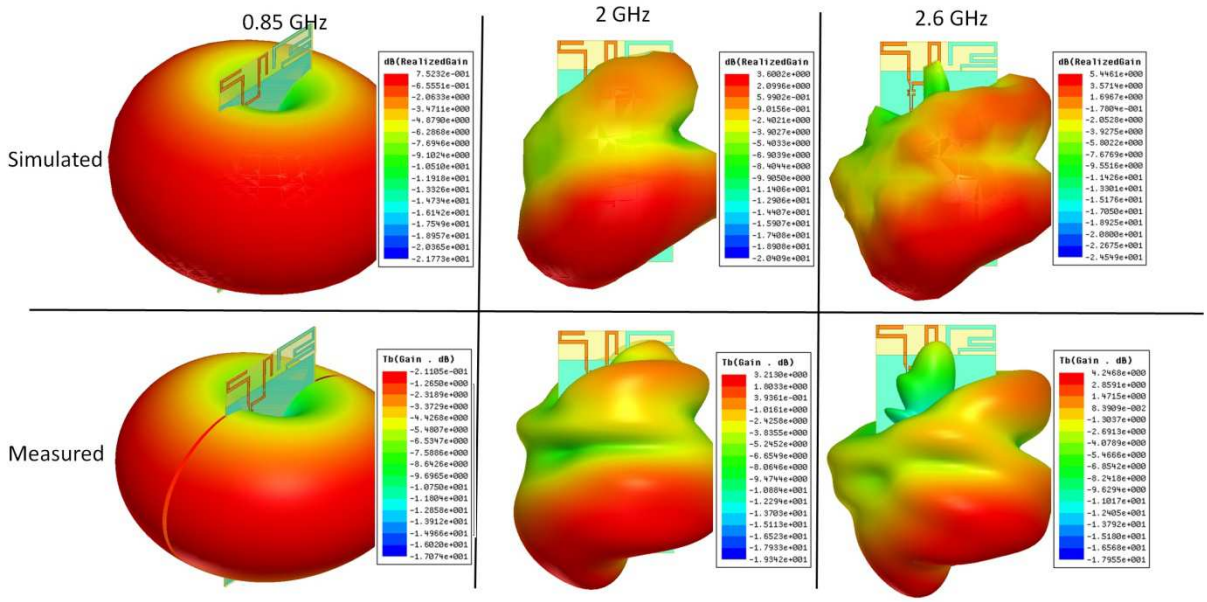

Figure 48 Simulated and Measured 3D Gain Patterns for Antenna Design II

A radiation efficiency drop issue can be also observed also for this prototype around 1.8GHz in Figure 49. It is worthwhile to note that the measured radiation efficiency in this figure is calculated from the measured total efficiency using the measured reflection coefficient. The efficiency drop is observed again at a frequency point, where the parasitic strip is strongly excited. Explanations about this problem will be given with some proposed solutions in next Section 1.3.

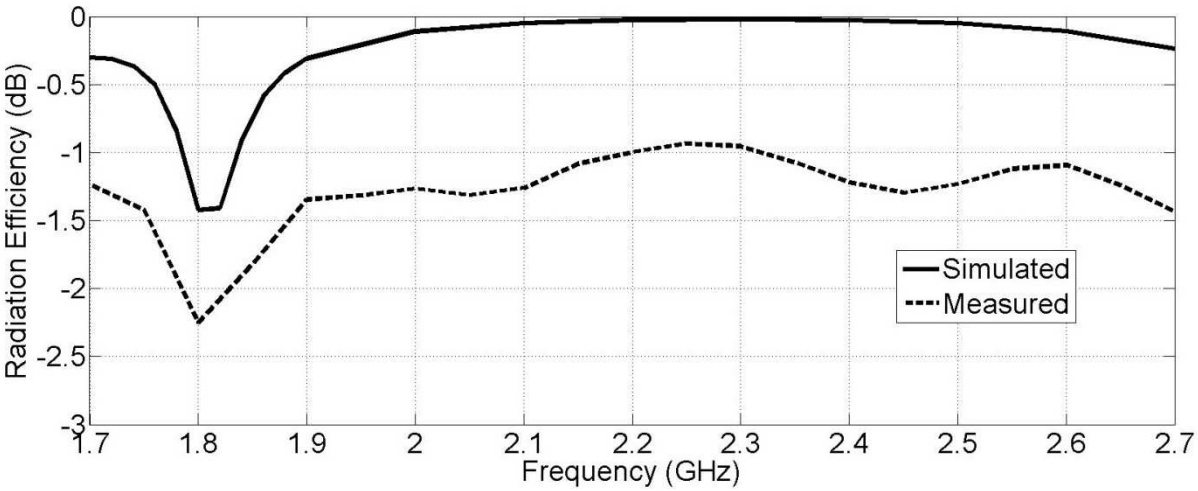

Figure 49 Simulated and Measured Radiation Efficiency in HB for Antenna Design II

1.3. RADIATION EFFICIENCY DROP IN ANTENNA PROTOTYPES

As presented in Figure 40 and Figure 49 of this chapter, a narrowband drop in the radiation efficiency was observed for both prototypes both in simulation and measurement, at the frequency regions corresponding to the excitation of the parasitic element in the HB. For the situation here, the drops observed are not causing a deep null in the radiation efficiency, but instead decreasing it as much as between 0.5-2dB from the general trend in simulations. However it should be noted that depending on the antenna topology, the problem may be more dramatic causing deep efficiency drops.

To investigate the reason of this situation, the surface currents were plotted in Figure 50 for antenna design I, at the frequency points of 1.86GHz and 2.46GHz, where a radiation efficiency drop of approximately 10% (-0.5dB) is observed from the general trend of the neighboring frequencies. Since the parasitic strip contributes to the HB frequency behavior at these two frequencies, strong currents can be observed on it (encircled with black ellipses in Figure 50). The direction of the current flow stays in opposite directions around the meander region, in a physically close manner. The fields created from these currents tend to cancel each other in the far-field region of the antenna, which leads to a radiation efficiency drop.

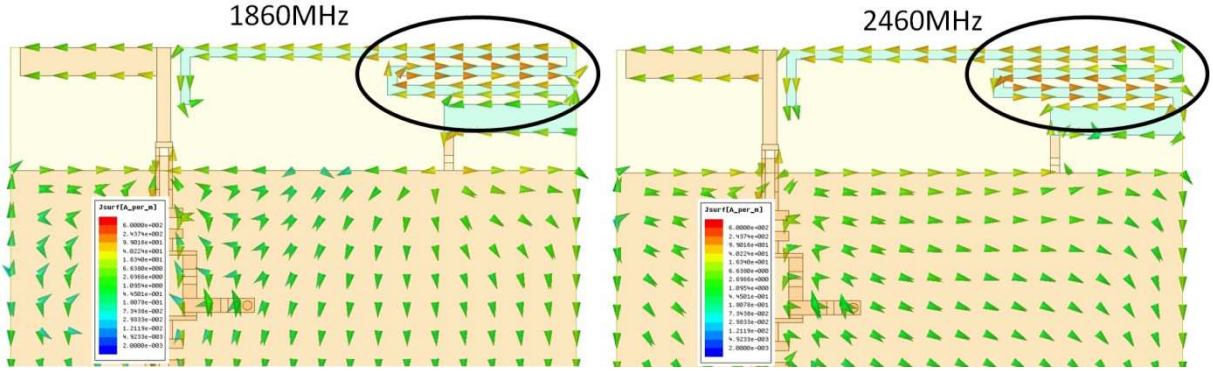

Figure 50 Surface Currents for the Efficiency Drop in Antenna Design I

To mitigate this effect and also for verifying the cause of the drop (suspected as the reverse currents on the meander region), the topology of the parasitic element was modified as seen in Figure 51, removing the meander line. Since the electrical length of the parasitic element is now shorter, the SMD inductor value on the ground connection is increased from 12nH to 16nH. In this way, the role of the parasitic element in the LB is maintained (for creating an additional resonance to increase bandwidth potential). For the simulations to be fair, no other property of the antennas nor the component values were changed. It can be seen in Figure 51 with the simulated reflection coefficient that the target bands can still be covered after this modification, although some tuning might be needed in the HB. The radiation efficiency of the modified structure is compared with the original case in Figure 52. The drops in the radiation efficiency are now satisfactorily solved with the modified structure.

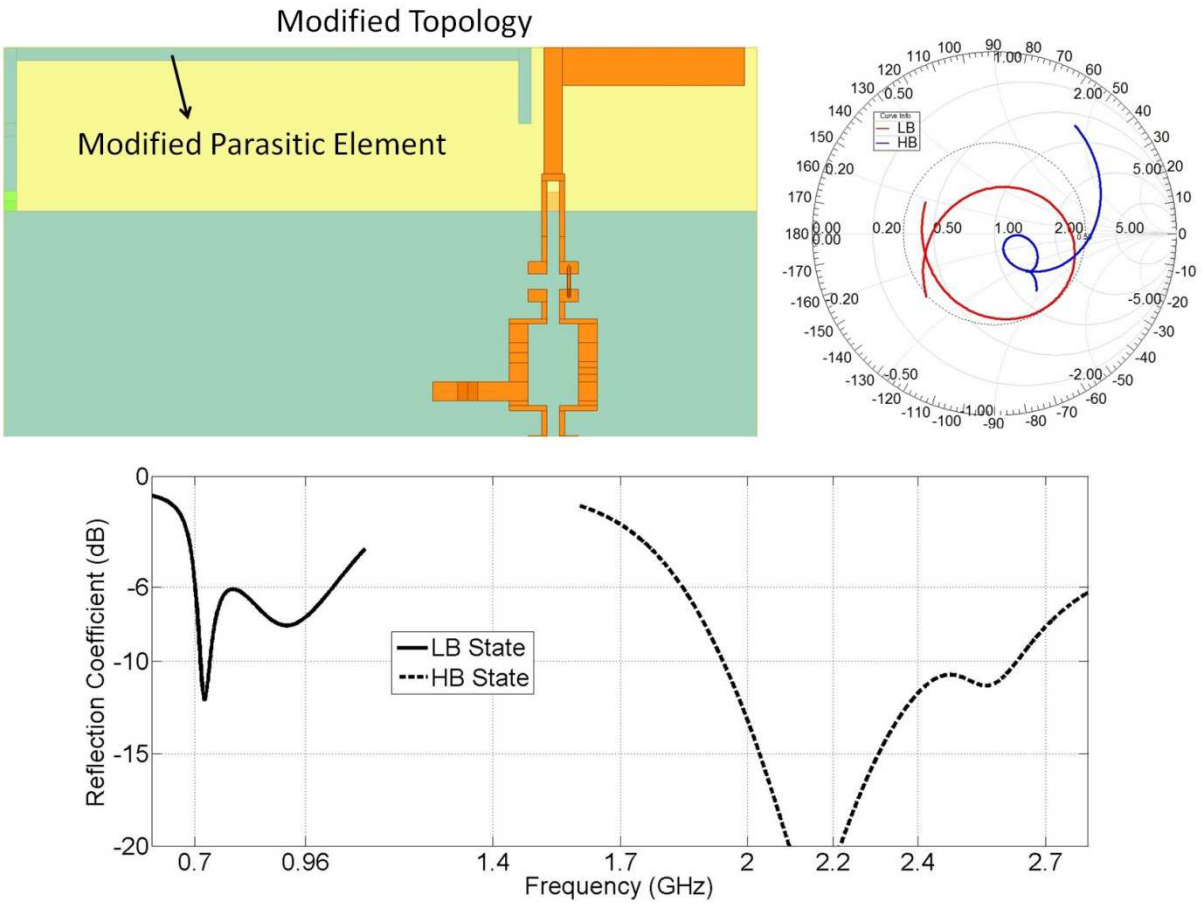

Figure 51 Modified Topology to Mitigate the Efficiency Drop for Antenna Design I and Simulated Input Impedance and Reflection Coefficient

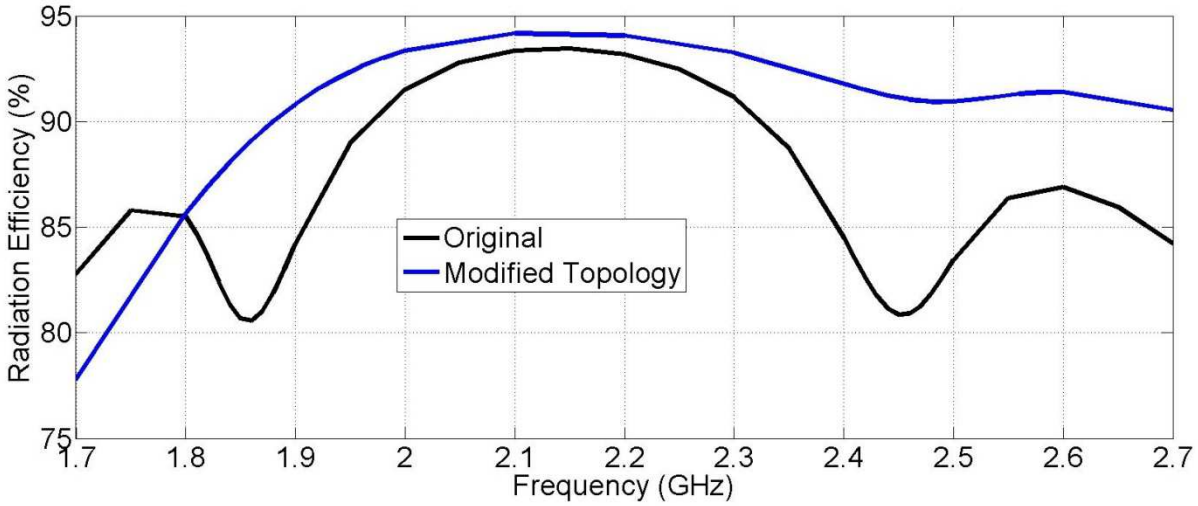

Figure 52 Simulated Radiation Efficiency of the Modified Topology for Antenna Design I

A similar study was done with antenna design II also. The surface currents in opposite directions on the meander line can be observed in Figure 53 at 1800MHz.

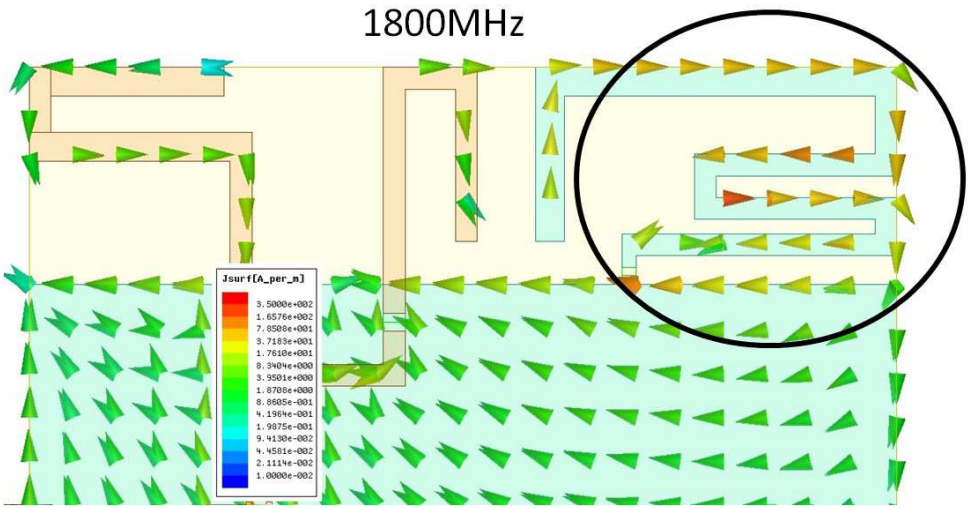

Figure 53 Surface Currents for the Efficiency Drop in Antenna Design II

Two different modified topologies are proposed in Figure 54, avoiding any meander line on the parasitic strip. To compensate the shortening of the electrical length of the parasitic element, the SMD inductor value on the ground connection point was increased to 27nH in modified topology-1 and to 29nH in modified topology-2 (from 18nH). A good matching behavior concerning the target frequency bands can be achieved with the two topologies.

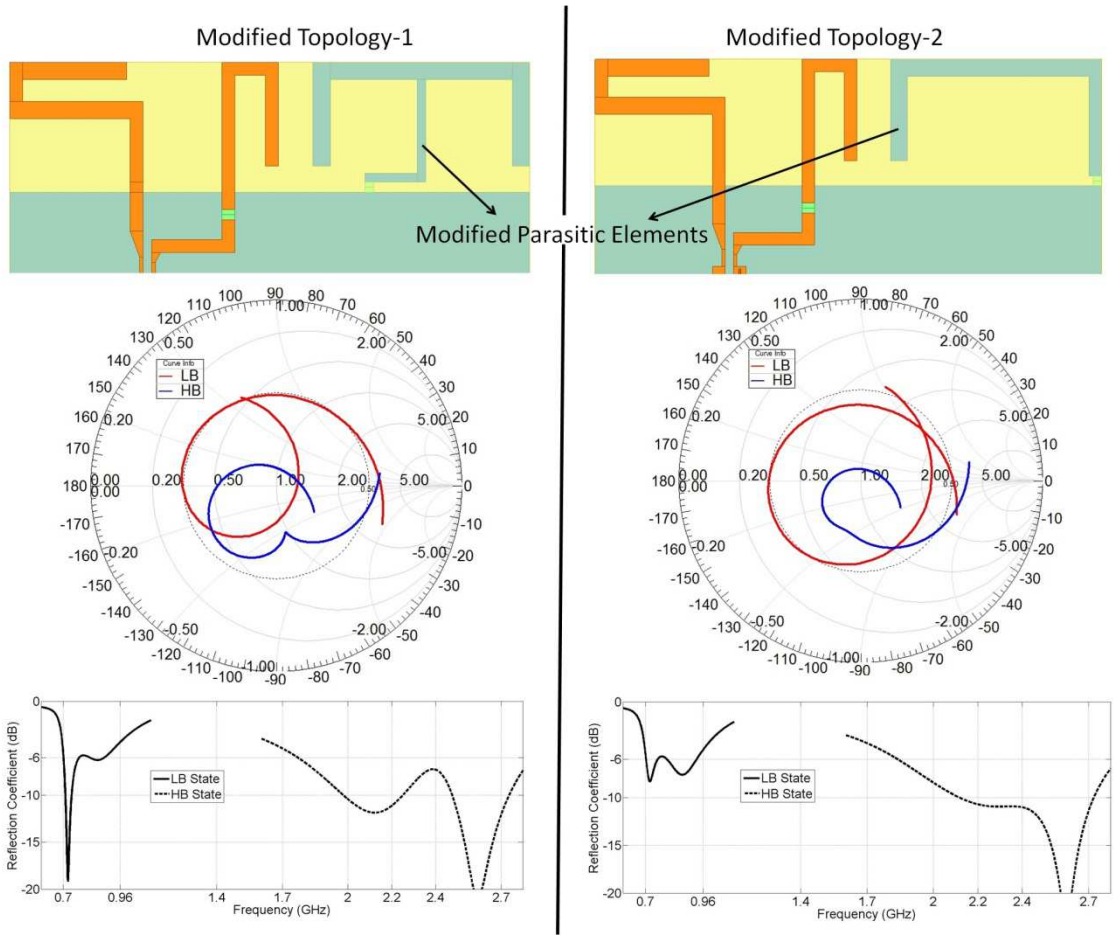

Figure 54 Modified Topology to Mitigate the Efficiency Drop for Antenna Design II

The effect of this modification on the simulated radiation efficiency is presented in Figure 55. Although the 25% efficiency drop at 1.8GHz is prevented, a drop of approximately 10% is again seen in modified topology-1 at the frequency where the parasitic strip is excited. This can further be avoided by modifying the layout as in modified topology-2.

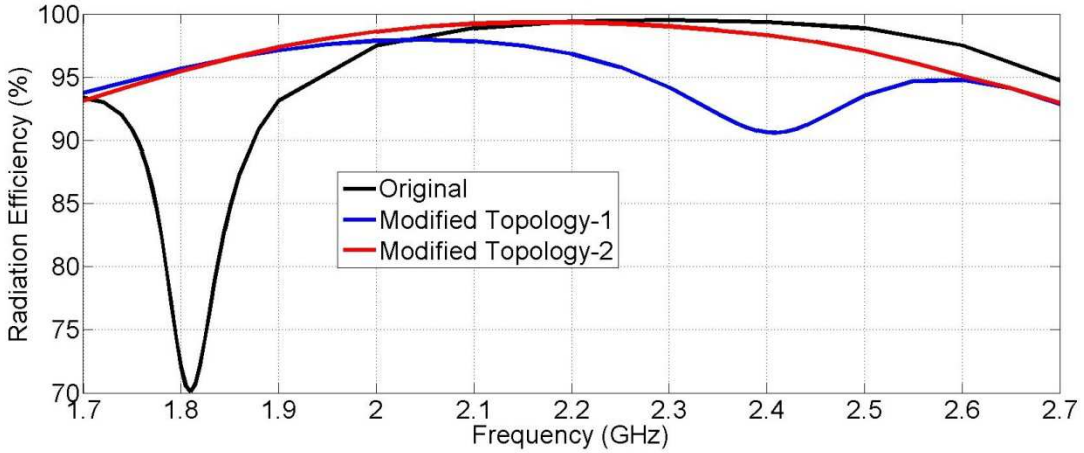

Figure 55 Simulated Radiation Efficiency of the Modified Topology for Antenna Design II

The fabrication of the modified antennas were not possible due to time constraints so the measurement results are not presented here.

2. ANTENNA DESIGN USING BST CAPACITOR

Another tunable component for use in the tunable antennas from the Artemos Project is the tunable BST(**B**arium **S**trontium **T**itanate) capacitors from CEA-LETI[12]. This type of tunable capacitors have been used for matching band reconfigurability in [13-14]. In [13], a tunable capacitor with a tuning range between 1.4-4.2pF was used in a tunable matching network at the feed of a COTS chip antenna to tune the matching bandwidth between 1.55-1.7GHz(Figure 56). A tunable dual-band IFA was proposed in [14], whose matching bands are tuned using a series LC resonator in parallel with a BST capacitor on the IFA. In this way the LB of the antenna is tuned between 822-1050MHz and HB is tuned between 1.42-2.2GHz using the BST capacitor having a tuning range of 3.3:1.

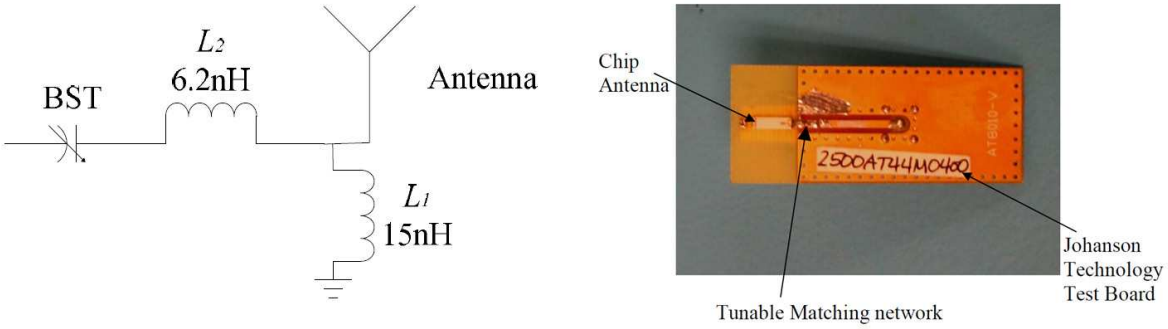

Figure 56 Tunable Matching Network and Test Prototype in [13]

For this study, CEA-LETI has implemented tunable Metal/Insulator/Metal capacitors using $Ba_{0.7}Sr_{0.3}TiO_3$ as the dielectric material. A general layout of the BST tunable capacitor can be seen in Figure 57. Tunability in the capacitance value is achieved by changing the applied bias voltage between the top and bottom metal plates. This voltage difference changes the permittivity of the BST material, which in turn changes the capacitance between the plates.

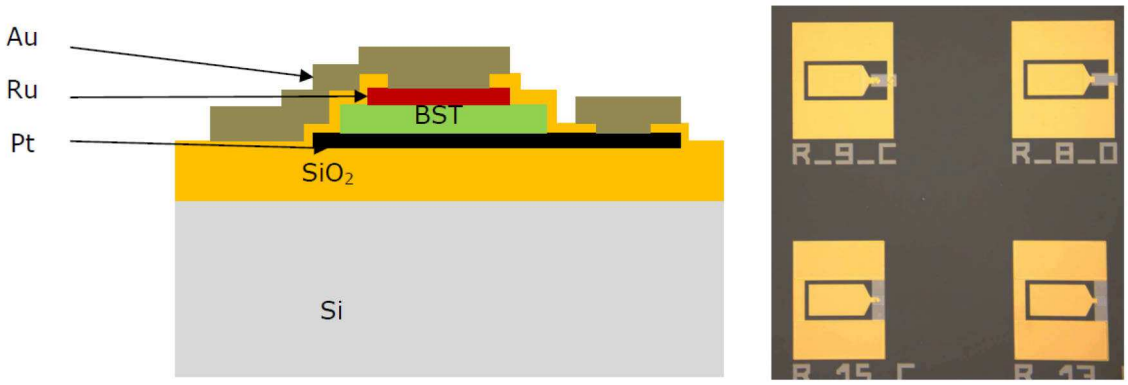

Figure 57 Layout of a BST Capacitor and Photograph of Manufactured Dies

The main challenge of using these tunable capacitors is the relatively low tunability range, which was agreed to be between 2.5-4pF with the manufacturer. The measured capacitance value versus the applied bias voltage for the BST capacitor can be seen in Figure 58. The tunable capacitor is exploited to achieve matching tunability in the LB, with a simple and low-cost antenna geometry. For the coverage of the HB, another antenna which is separately fed is used, obtaining a dual-feed configuration.

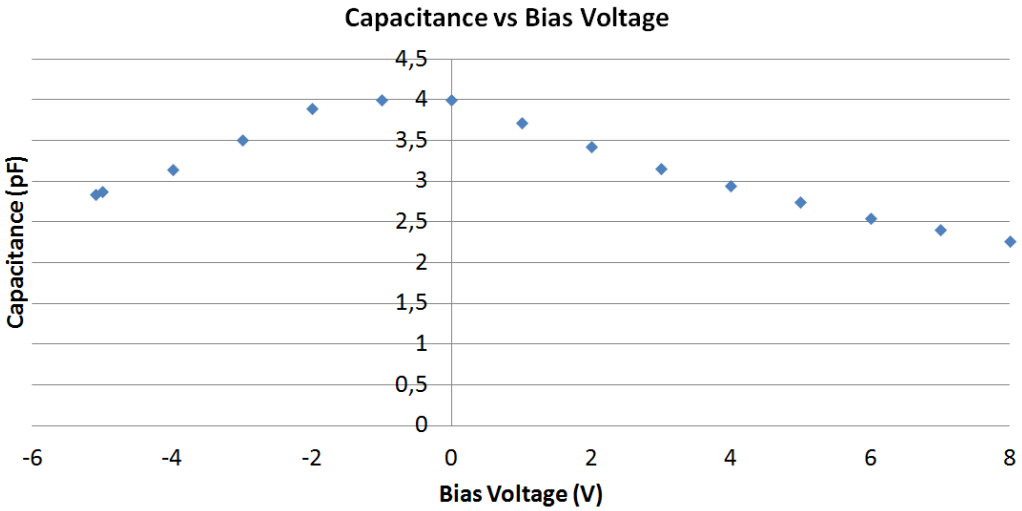

Figure 58 Measured Capacitance vs Bias Voltage for BST Capacitors

2.1. ANTENNA MODEL AND SIMULATION RESULTS

The proposed antenna model can be seen in Figure 59. As already mentioned, the antenna has a dual-feed structure, where one of the feeds is responsible for LB coverage and the other is responsible for HB coverage. All the antenna structure is again printed on FR4 substrate, having total dimensions of 120mm X 59mm X 0.8mm, with a ground clearance of 15mm on one of the short edges. The LB antenna is a simple coupling element (CE), which is tuned to the desired band using a pre-matching network and the tunable BST capacitor. The HB antenna is a simple monopole, fed through the second feed of the antenna.

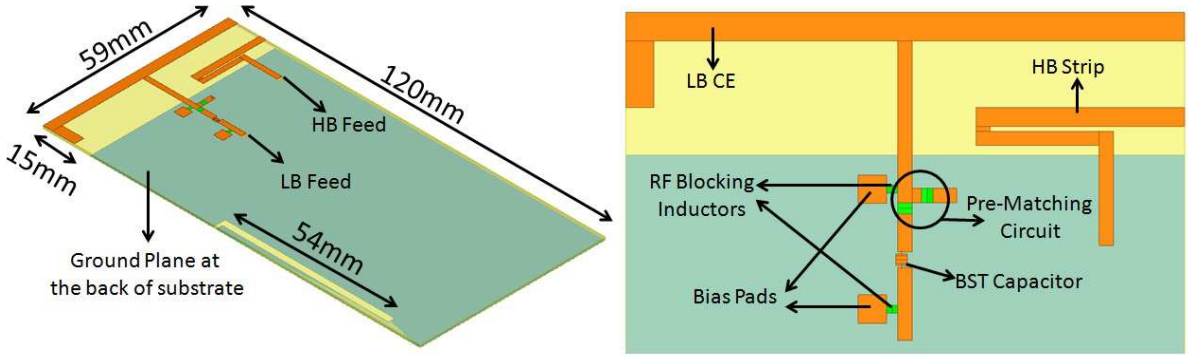

Figure 59 Simulation Model of the BST Antenna

Since the tunable capacitor has a relatively low tuning range (4 : 2.5), a slot of 54mm length was etched on the ground plane, to create an additional resonance around 950MHz, which makes the coverage of the LB easier. The slot basically changes the current distribution on the ground plane and the chassis resonance shifts to lower frequencies since the electrical length of the ground plane increases. The length of the slot is an important parameter in the design phase of the antenna. Changing this length modifies the frequency of resonance, seen in the blue curve in Figure 60. The length was optimized as 54mm after a parametric study. The effect of the slot can be observed in Figure 60, plotted between 700-960MHz.

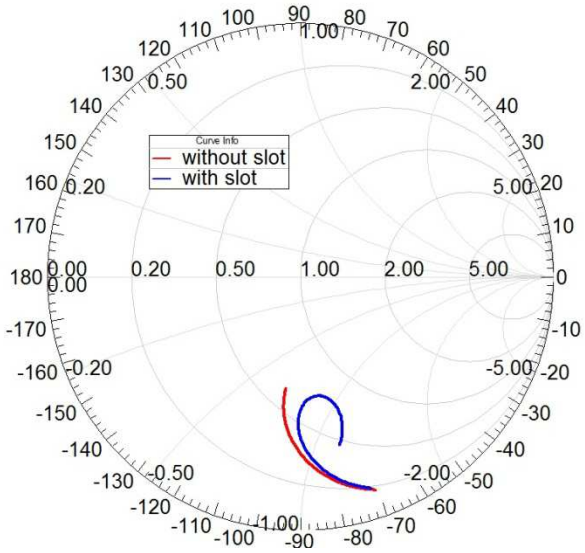

Figure 60 Effect of the Slot in the Ground Plane

The left smith chart in Figure 61 presents the simulated input impedance of the antenna when the LB CE is simulated in the absence of the HB monopole and vice versa. The red curve is plotted for frequency interval 700-960MHz and blue curve plotted between 1.7-2.7GHz. The response of the combined antenna structure, after the pre-matching using a shunt inductor of 30nH and series inductor of 22nH is given on the right in Figure 61. One thing to notice here is the additional resonance created in the HB around 1900MHz. According to the surface currents plotted in Figure 62, this new resonance comes from the currents induced on the LB CE. Afterwards these currents flow to the ground plane over the shunt inductor and also to the LB feed of the antenna, which should be kept under a specific level to achieve enough port-to-port isolation.

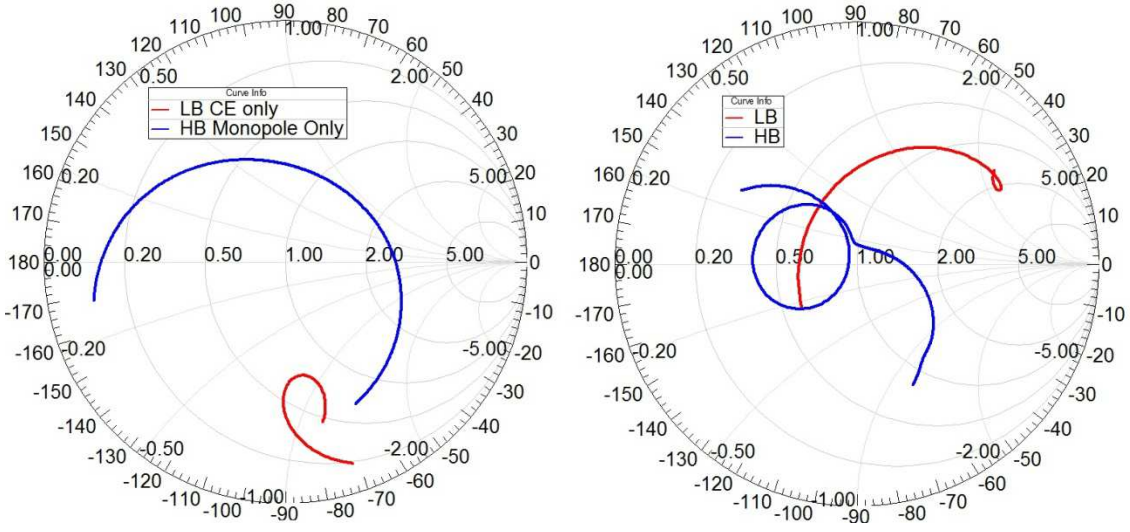

Figure 61 Simulated Input Impedance of the Antenna (LB CE and HB Monopole Simulated Alone on left, After Pre-Matching on right)

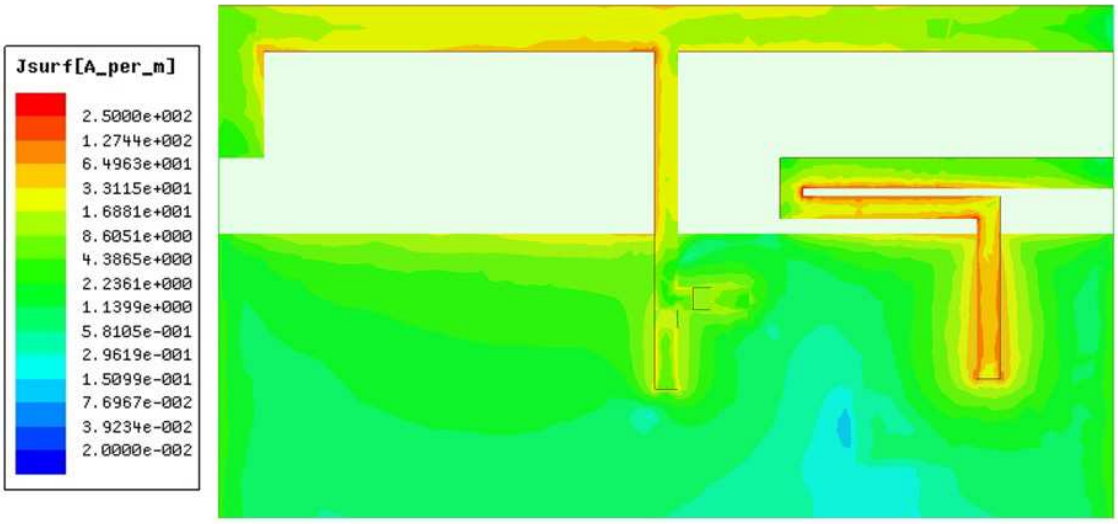

Figure 62 Surface Currents for BST Antenna at 1900MHz After Pre-Matching

Simulations were done for the antenna taking into account the wirebondings and RF Blocking inductors (120nH) for the DC bias pads, that are necessary for the integration of the BST capacitor. For the SMD components, the measurement results from the manufacturers were used. The simulated reflection coefficient in the LB for the BST capacitance changing from 2.5pF to 4pF is given in Figure 63. By increasing the capacitance value, the impedance curve moves in clockwise direction on a constant resistance circle in smith chart, enabling the coverage of lower frequencies. The resonance created by the slot in the ground plane helps considerably for the matching in the high-edge of the band. With this configuration, the frequency interval of 725-960MHz can be covered with a reflection coefficient below -6dB.

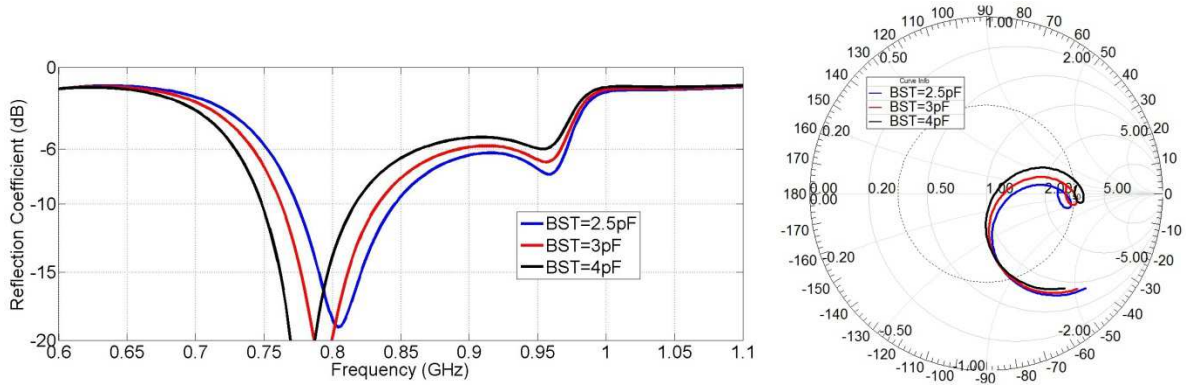

Figure 63 Simulated Reflection Coefficient in the LB for Different BST Capacitance Values

It was seen that the HB performance is not affected by the change in the tunable capacitor. Figure 64 shows the total antenna s-parameters when the BST capacitor is set to 2.5pF. The HB antenna can cover the 1.7-2.7GHz frequency interval with a reflection coefficient below -6dB. The isolation between the two ports is almost always higher than 10dB, making a minimum of 10dB around 2GHz. As shown in Figure 62, this frequency region is where the strong coupling to the LB CE from the HB monopole is observed.

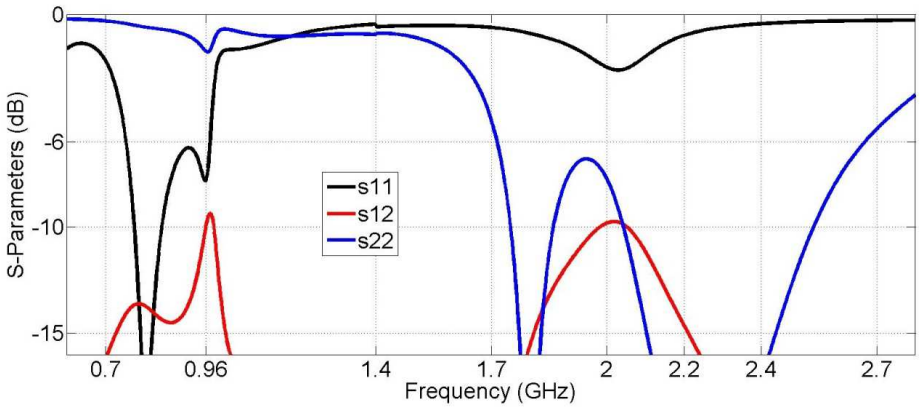

Figure 64 Simulated Reflection Coefficient for BST=2.5pF

The proposed antenna is currently under the manufacturing phase, which will be followed by the integration of the BST capacitor and the measurements to validate the simulation results.

3. CONCLUSIONS

In this chapter, three tunable antenna topologies have been proposed. Two of these antennas use MEMS switches to achieve band-switching between the LB (700-960MHz) and HB (1700-2700MHz), whereas the third antenna uses a BST tunable capacitor to tune the LB reflection coefficient.

The main difference between the two antenna using MEMS switches was the methodology of utilizing the switches. The first antenna uses two MEMS, to select between one of the MN branches and the second antenna uses a single MEMS to select between two driven strips. The simulation results have been validated with the measurements using manufactured passive prototypes for the first design and the antenna was shown to cover the target bands with a reflection coefficient below -6dB. For the second design, measurement results with both passive prototypes and real MEMS switches were presented.

In Section 1.3, a situation observed during the efficiency measurements of the antennas was investigated. A drop in the radiation efficiency was seen in both simulations and measurements, which is narrow-band and corresponds to the frequency regions where the parasitic element is excited capacitively in the HB. It was shown through simulated current distributions that the cause of this situation was the currents flowing in opposite directions on the meander line, which tend to create cancelling field in the far-field region of the antenna. Some modifications on the antenna topologies were successfully proposed to mitigate this effect.

Finally, an antenna design using a BST tunable capacitor was presented, that achieves tunability in the LB reflection coefficient. This antenna was a dual-feed structure where the HB antenna is separately-fed and is not affected by the changing capacitance value. It was shown through simulations that the frequency interval of 725-960MHz and 1.7-2.7GHz can be covered with a tunable capacitance ranging between 2.5-4pF, with an isolation higher than 10dB between the feeds.

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CHAPTER III
THE HOLLOW COUPLING ELEMENT APPROACH

As previously mentioned in the Introduction part, coupling elements (CE) have been widely proposed in literature for mobile terminal antennas, to obtain a wide-band matching response using the space available for the antenna, which is typically electrically-small. In this scenario, the CE is used just as an exciter to induce the proper currents on the terminal ground plane, whose dimensions are generally close to a 0.25-0.3 free space wavelength in the frequency band of 700-960MHz. In this way, the quality factor of the antenna is significantly reduced thanks to the low-Q nature of the thick ground plane and a high bandwidth potential is enabled. Since they are inherently non-resonant, CEs need a matching network at the antenna feed, to be tuned to the desired operating frequency. The main drawback of using CEs for 4G/LTE standards is the necessity of MNs with a high number of components to achieve dual-band coverage which increases both the losses and the cost.

This chapter presents a new design, namely a modified version of the traditional plain coupling element approach, by removing the inner metal part of the CE. This way, a hollow structure is obtained and the space available inside the CE can be used to insert another antenna to cover other frequency bands. In terms of input impedance, negligible difference is observed in the LB, by making the CE hollow. Two antenna designs using the new approach are designed, optimized and presented to cover the 700-960MHz and 1.7-2.7GHz bands. Finally, a dual-antenna MIMO structure using the same technique is presented as an application example to operate in the LB frequency range.

1. THE HOLLOW COUPLING ELEMENT APPROACH

The proposed approach combining a non-resonant CE and a resonant antenna is presented in this section. The LB coverage is achieved using a 3D CE. The HB coverage is then achieved by integrating another antenna with a separate feed.

1.1. COMPARISON OF THE HOLLOW CE WITH A PLAIN CE

As previously mentioned, the traditional plain CE has been modified to obtain a more space-efficient structure with negligible performance trade-off. The main issue is to produce the same coupling behavior as the one produced with a plain element or more generally, to understand which section of a plain CE is important to couple to the PCB. This approach was first described for a single band CE with a 7mm height, covering only the LB 700-960MHz frequency range in the previous publications of the authors [1]. In [2], the HB antenna was inserted inside the hollow CE and the dual band coverage with good isolation was presented with simulations. Here, a further optimized CE layout in terms of occupied space (only 3mm height) is used to make a comparison between plain and hollow CEs in terms of impedance and bandwidth potential. For the comparison, the model given in Figure 65 have been used.

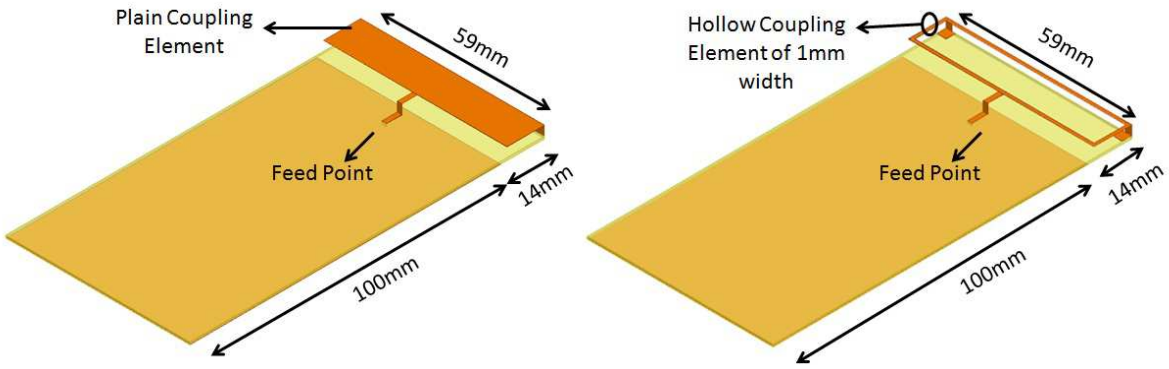

Figure 65 Antenna topologies used for the plain/hollow CE comparison

The plain CE consists of a rectangular metal sheet (59×14mm2) with 3mm height from the FR4 substrate. The feed of the CE is achieved through a vertical metal strip from the center of the PCB. The total PCB dimensions are 114×59×0.8mm^3 with a ground clearance of 14mm under the CE. For a fair comparison, all PCB dimensions and feeding location were kept the same for the hollow. The metal region inside the plain CE was removed to form the so-called hollow CE, leaving only a rectangular ring having a strip width of 1mm. The two CEs were simulated with the full wave electromagnetic simulator Ansoft HFSS. The simulated input impedance of the two CEs can be observed in Figure 66. The input impedance of the hollow CE shows minor deviation from the input impedance of the plain CE.

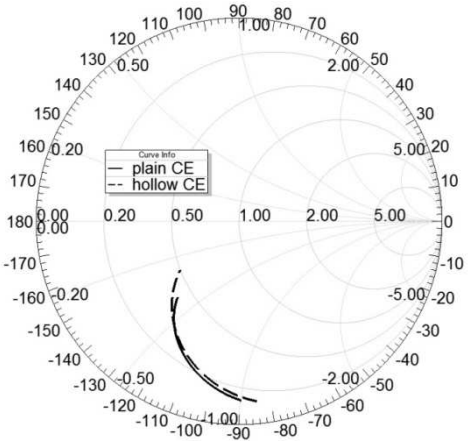

Figure 66 Input impedance of the plain and the hollow CEs (700MHz to 960MHz)

The currents induced on the system ground plane for the plain and hollow CEs can be observed in Figure 67. In both situations, the longitudinal mode (the fundamental) of the currents is induced almost at the same strength.

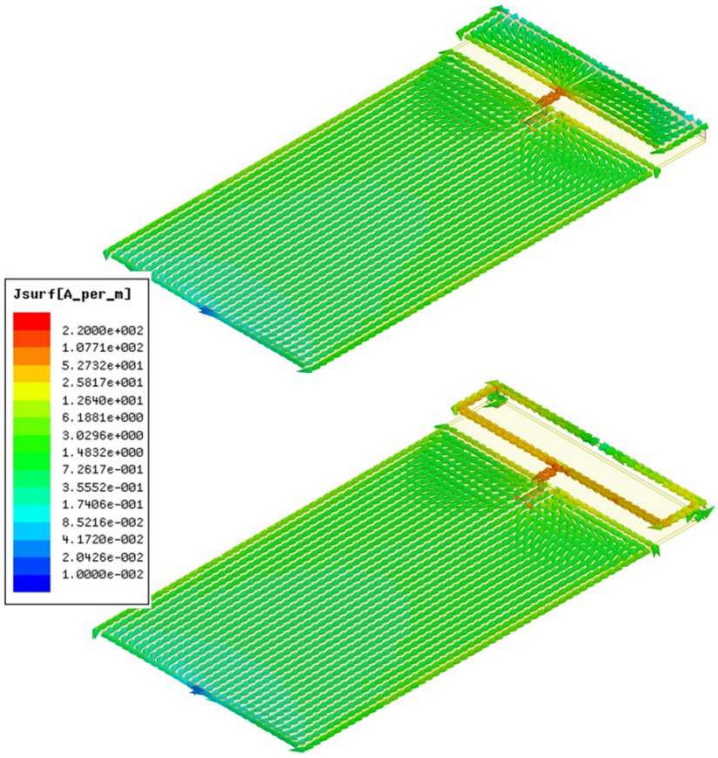

Figure 67 Comparison of the Ground Plane Currents (Plain CE on top, Hollow CE on bottom)

The E-field distribution under both CEs is observed in Figure 68. It can be told that fields under the hollow CE are weaker than the fields under the plain CE especially through the center of the hollow region. Using this fact, the HB antenna is proposed to be placed in this region, to get better port-to-port isolation.

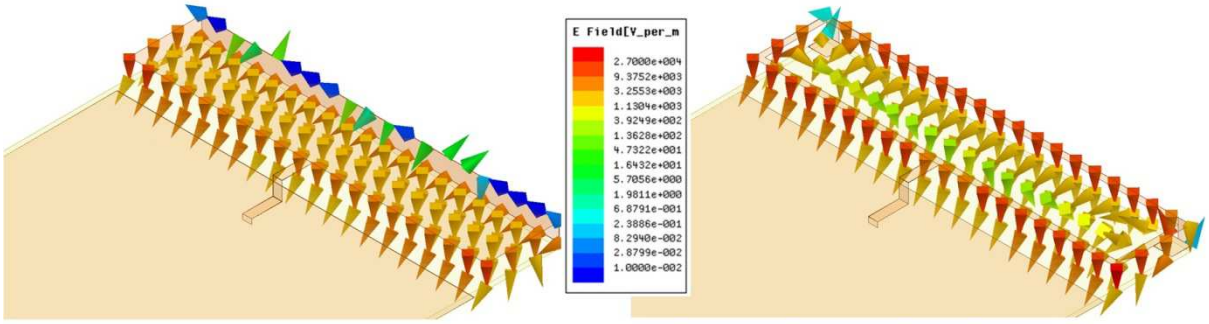

Figure 68 Comparison of the E-field Under the CEs (Plain CE on left, Hollow CE on right)

The bandwidth potential (BP) of the two CEs is compared with the help of the commercial software Optenni Lab. The BP is a realistic value computed from the input impedance of each CE (obtained from the full wave simulation) which is given for a certain number of matching components and a certain threshold for the reflection coefficient. At every frequency point f_0, the software generates a MN to match as much as possible frequency points around f_0 with a reflection coefficient below the threshold, chosen as -6dB here. The frequency interval around f_0 which gives a reflection coefficient below -6dB is given as the optimized BP (symmetrically around f_0). The BP obtained in this way is a realistic value compared to very optimistic theoretical values computed from the quality factor. As can be seen in Figure 69, the BP of the hollow CE is only slightly lower

(30MHz) than the BP of the plain CE in the LB, which proves again the equivalence between the plain and the novel hollow CE in terms of matching behavior. In the center of the 700-960MHz band, the obtainable BP is around 200MHz (for a -6dB reflection coefficient). It should be kept in mind that this BP value is obtained with a MN of two lumped components which is foreseen to be increased above the required 260MHz to cover the full LB by using more matching network components.

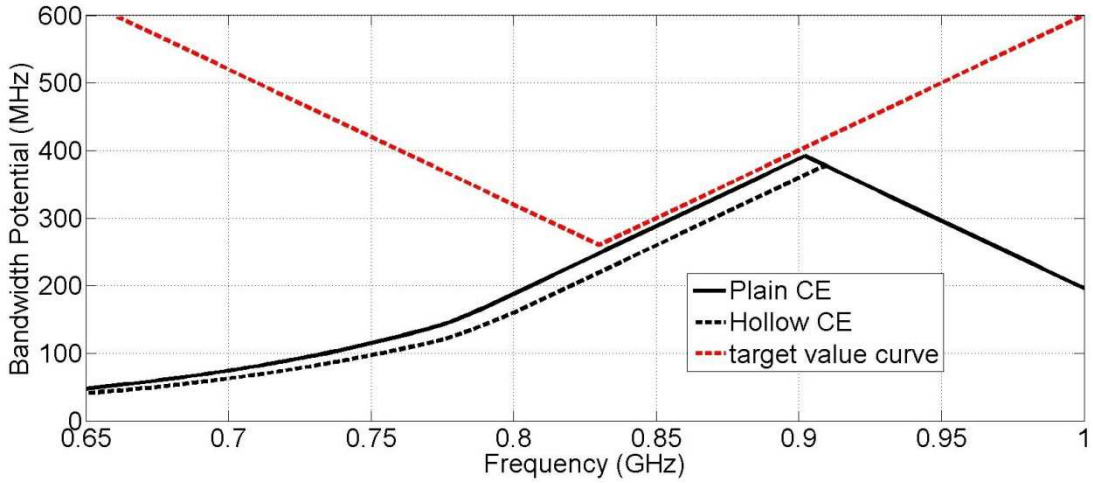

Figure 69 Bandwidth Potential Comparison of the Plain and Hollow CEs

1.2. INTEGRATION OF HIGH BAND ANTENNA WITHIN THE HOLLOW CE

As previously mentioned, the space gained by making the plain CE hollow can now be used to integrate a high-band antenna. For ease of manufacturing, it was decided to focus on a fully printed HB antenna design (directly on the FR4 substrate). Since it was observed that the E-field under the hollow CE was weaker through the center of the hollow region (Figure 68), the HB antenna was planned to be placed in this area to obtain high port-to-port isolation.

The optimized layout of the proposed antenna can be seen in Figure 70. The total PCB dimensions are 114×59×0.8mm^3, with a ground clearance region of 59×14mm^2 on one side of the PCB. A simple monopole is used for the HB coverage, placed 8.5mm away from the edge of the PCB. To tune the hollow CE as to cover 700-960MHz, a MN (seen in Figure 71) consisting of four lumped components was designed with the Optenni Lab software. The main objective of this MN is to get a reflection coefficient below -6dB for the hollow CE between 700-960MHz. There is also another functionality of the MN, which is to act as a band-stop filter in the HB frequency range to help in resisting the possible currents from the HB antenna to reach the LB antenna port. Very high isolation levels were possible using this strategy in the HB, so the matching response of the HB antenna was the primary goal when determining the exact location and orientation of the HB antenna.

After the circuit optimization of the MN, an additional full-wave optimization was performed to take into account the existing transmission lines between the SMD components and the vias connected to the ground plane.

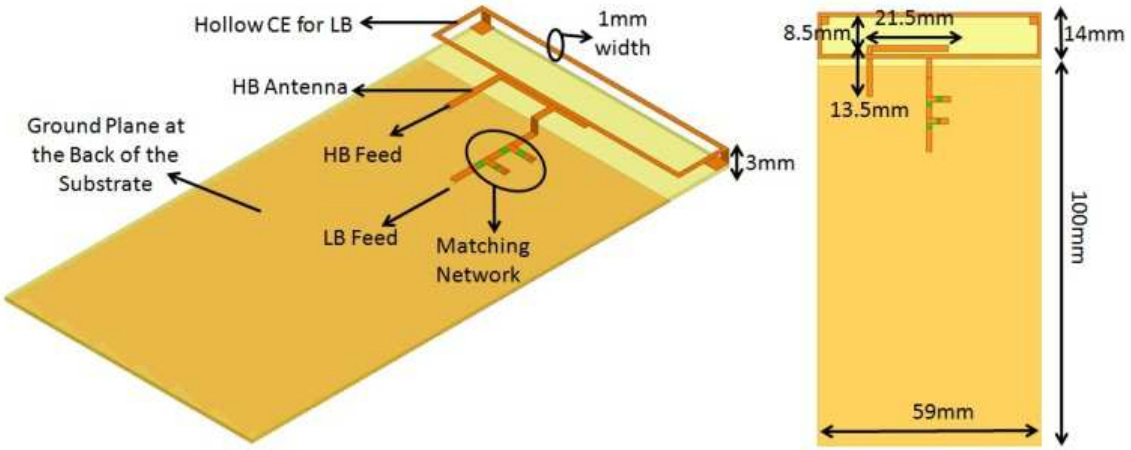

Figure 70 Simulation Model of the Hollow CE Antenna

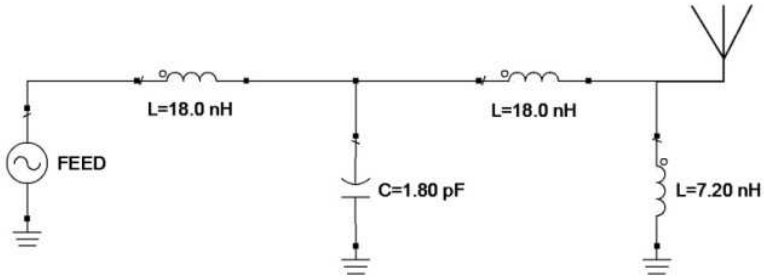

Figure 71 Matching Network used at the LB CE feed

The simulated input impedance of the LB antenna alone (without the HB antenna) with and without the MN can be seen in Figure 72, plotted from 700MHz to 960MHz. When the HB antenna is introduced, the S-parameters seen in Figure 73 are obtained. It can be seen that the target bands can be covered with a reflection coefficient below -6dB, except a small frequency range on the lower edge of the HB (1.7-1.75GHz). The port-to-port isolation is always higher than 9dB in the LB having a minimum value of 9.3dB at 740MHz. The port-to-port isolation is higher than 30dB in the HB.

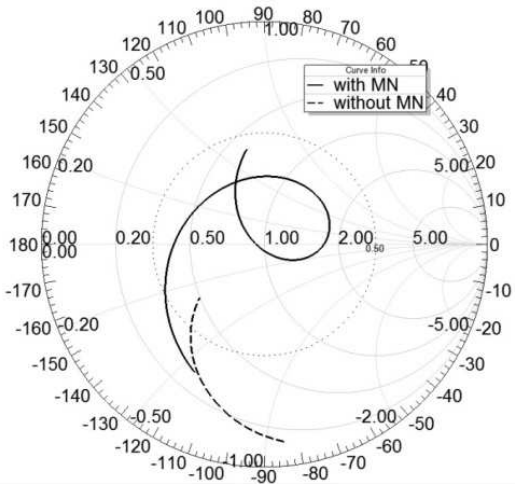

Figure 72 Simulated Input Impedance of the LB CE Alone (with and without MN)

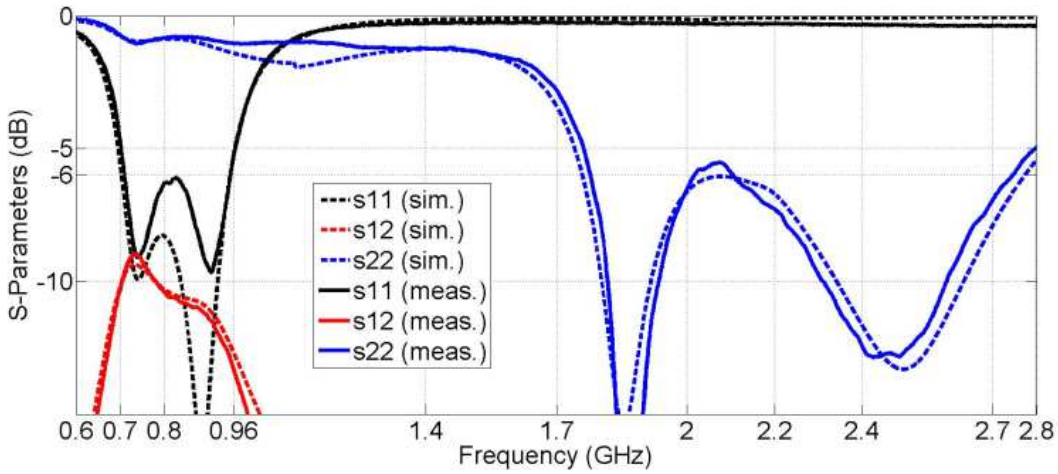

Figure 73 Simulated and Measured S-parameters of the Hollow CE Antenna

To investigate the role of the LB CE for the HB performance, the HB antenna was also simulated alone and with the CE. Figure 74 presents the simulated input impedance of the HB antenna (plotted between 1.7-2.7GHz) in these two cases. It is obvious that the wideband HB response is not due to the HB antenna alone. As can be seen from the input impedance shown in smith chart, an additional resonance in the HB is created by the coupling from the HB antenna to the LB CE. This coupling does not induce an isolation problem since the MN of the LB antenna is designed on purpose to provide a stop-band behavior from 1.7 to 2.7GHz. Proof is given in Figure 75, where the current distribution on the LB CE created by the HB antenna is observed at different frequencies. At 2GHz, it is seen that the currents on the HB antenna are weaker and the currents on the LB CE are stronger which suggests a high level of coupling. However, the current induced on the LB CE are shorted to the ground plane through the first shunt inductor (7.2nH) of the MN. In this way, the port-to-port isolation is kept high. If the currents at 2.3GHz are observed, it is seen that the HB antenna is strongly excited whereas the coupling to the LB CE is weaker.

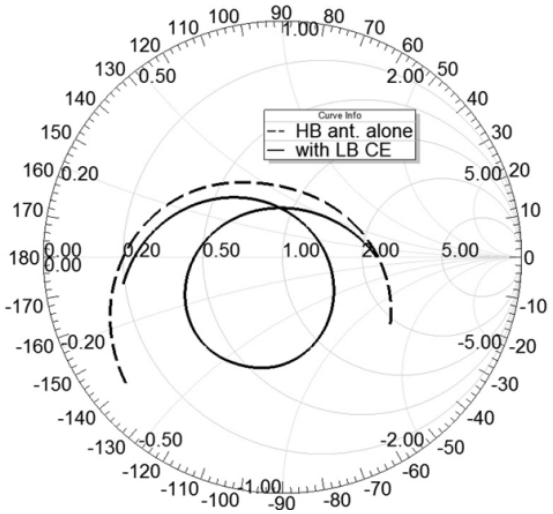

Figure 74 Simulated Input Impedance of the HB antenna alone and with the LB CE

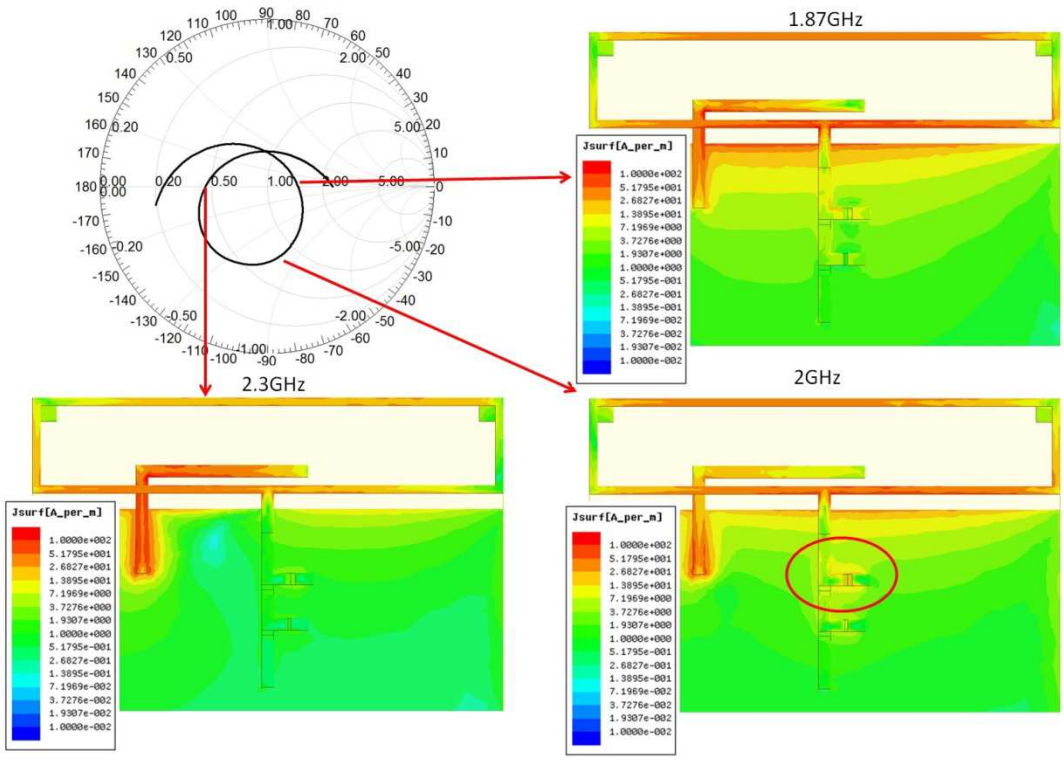

Figure 75 Current Distribution on the LB CE when the HB Port is Excited

1.2.1. Measurement Results

The proposed antenna was manufactured on FR4 substrate (Figure 76). As seen in Figure 73 and Figure 77, there is a fair agreement between simulated and measured S-parameters. The small differences between simulation and measurement are coming from the manufacturing tolerances in the 3D CE section and its orientation. The target bands are covered with a reflection coefficient below -6dB.

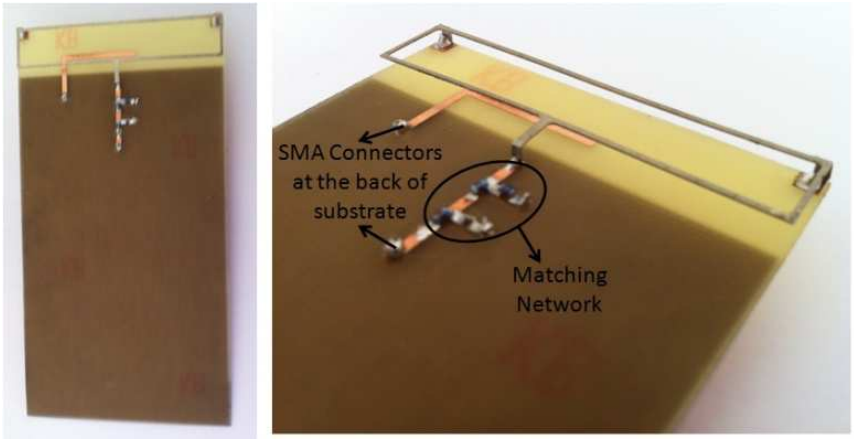

Figure 76 Manufactured Prototype of the Hollow CE Antenna

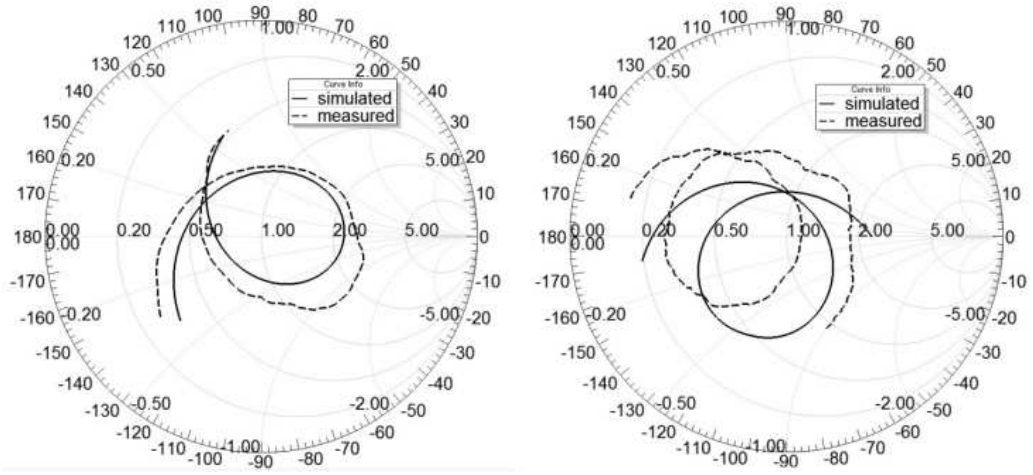

Figure 77 Simulated and Measured Input Impedance of the Hollow CE Antenna (LB on left, HB on right)

The total efficiency and the 3D gain patterns have also been measured in a Satimo Starlab Station. The total efficiency of the proposed antenna in LB and HB are presented in Figure 78 (the HB port is loaded by 50Ω when the LB efficiency is measured and vice versa). The measured total efficiency (including mismatch and component losses) in LB varies between -6dB and -2dB and between -4dB and -1dB in HB. When the simulated and measured 3D gain patterns are compared, a fair agreement is observed (Figure 79). In LB, a dipole-like radiation is observed (as expected) as the main radiator is the currents induced on the ground plane by the CE. Both in LB and HB, no wide-angle deep nulls are seen in the radiation patterns which is an important aspect for mobile phone communications where the base-station orientation is never known in advance.

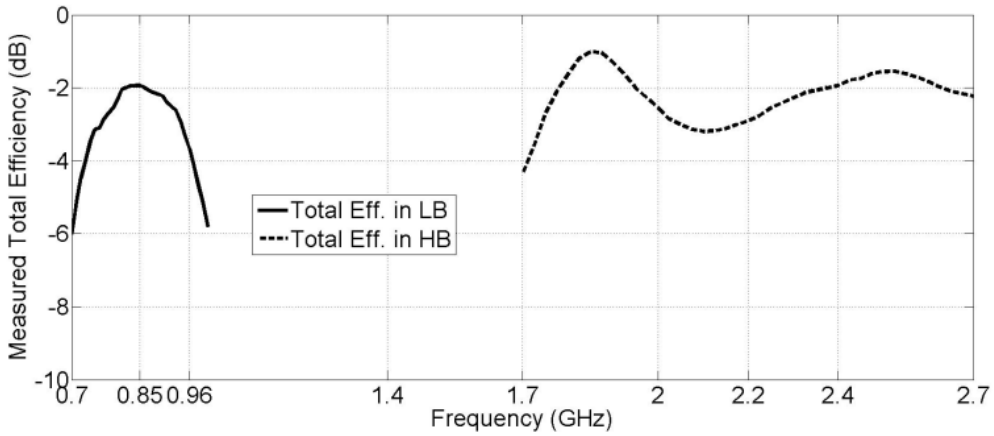

Figure 78 Measured Total Efficiency of the Hollow CE Antenna

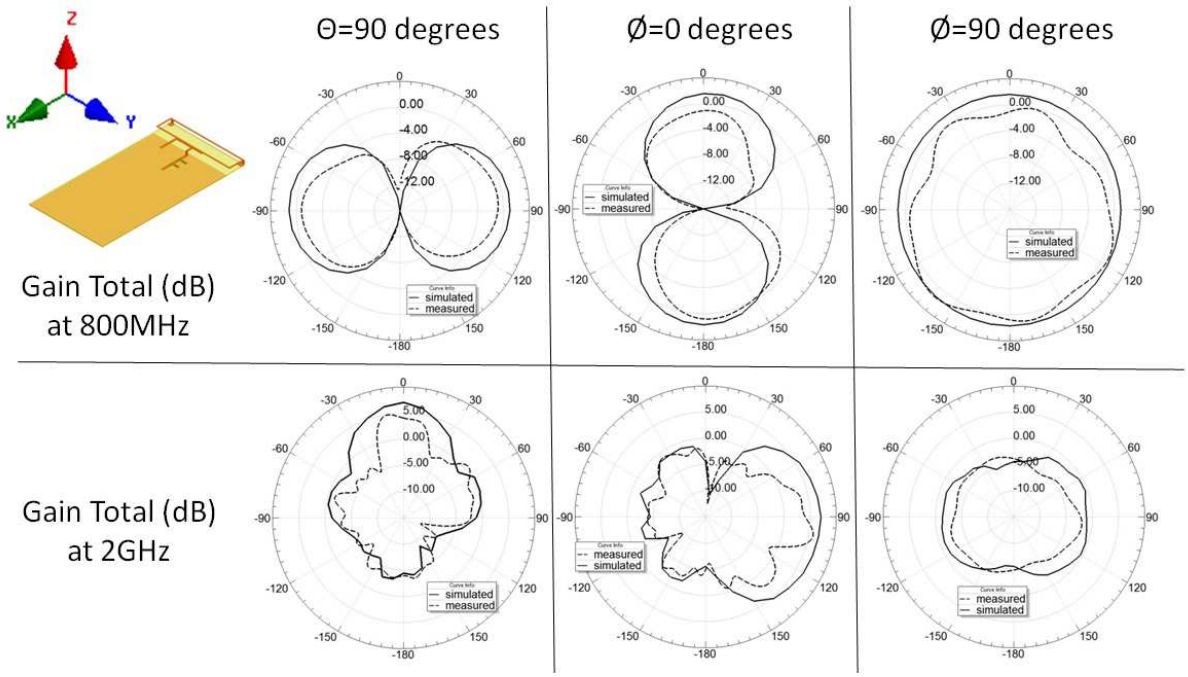

Figure 79 Simulated and Measured Total Gain Patterns for Different Cut Planes

1.3. PRINTED COUPLING ELEMENT ANTENNA

1.3.1. Antenna Design

The dual-feed antenna approach presented in Section 1.1 for a 3mm height 3D hollow CE was then enhanced to design an antenna-structure totally printed on the FR4 substrate for the ease of manufacturing and low cost purpose. The main difference with the previous antenna structure is the location of the printed HB antenna on the FR4 substrate. Since it was not possible to integrate the HB antenna inside the printed CE, it was placed on the left side of the PCB where the E-fields were found to be the weakest in simulations. The PCB and ground plane dimensions were set the same as the previous antenna structure (Figure 80). The dimensions of the hollow CE were set as 59×4mm2 with a strip width of 1mm. Moreover, a two-component MN was found to be necessary at the HB antenna port.

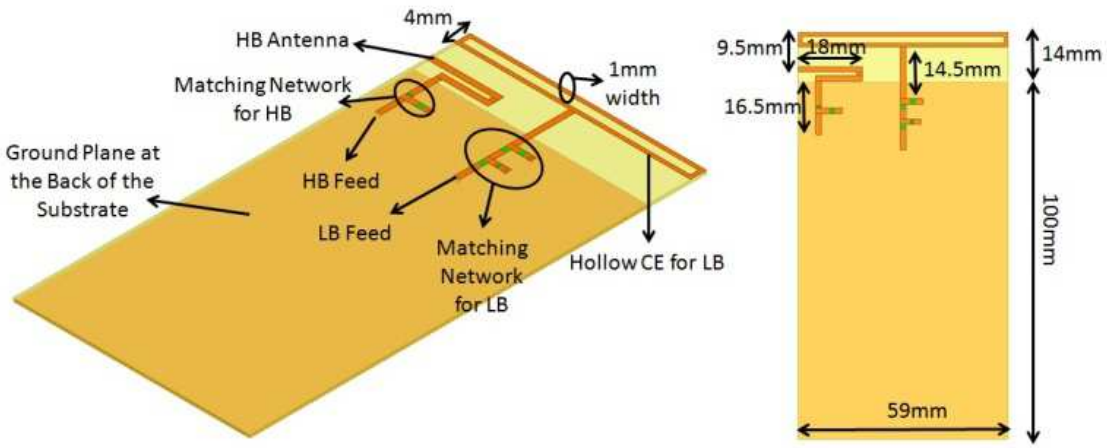

Figure 80 Simulation Model of the Printed CE Antenna

The effect of hollowing the printed CE instead of using a printed plain CE was investigated (Figure 81). Figure 82 shows a comparison of the bandwidth potential for the plain and hollow printed CEs, with the same dimensions. It was seen that making the CE hollow enables to cover an extra 25MHz around the center of the target LB. This is mainly due to the ring shape (loop) of the hollow CE which provides an inductive effect on its input impedance.

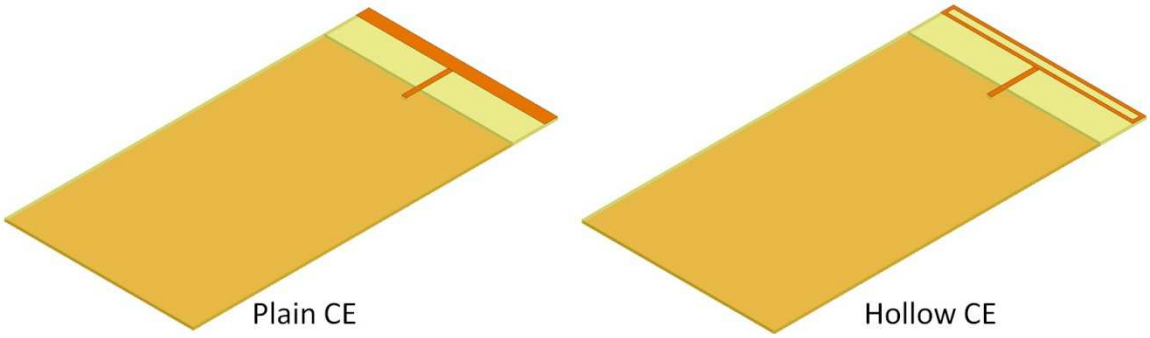

Figure 81 Plain and Hollow CE Geometries Used for Comparison

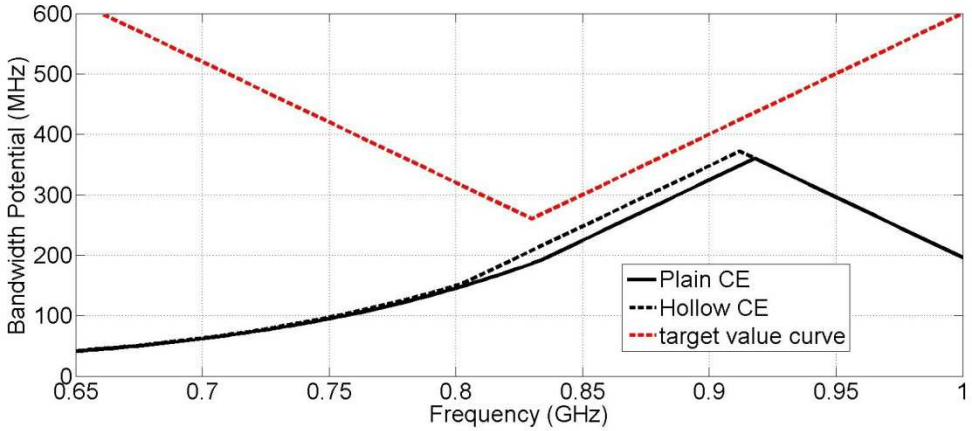

Figure 82 Comparison of Bandwidth Potentials for Plain and Hollow Printed CE

As seen in Figure 83, the real part of the impedance increases and the magnitude of the imaginary part decreases when the CE is made hollow. The e-field distribution between the hollow CE and the ground plane at 750MHz is plotted in Figure 84. The fields are the strongest close to the corners of the hollow CE and decreases as we move away from the CE.

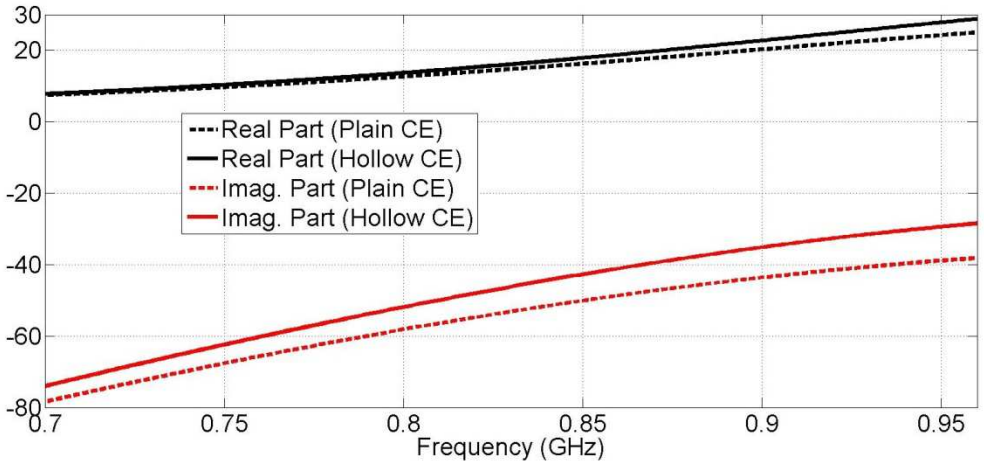

Figure 83 Comparison of Input Impedance for Plain and Hollow Printed CE

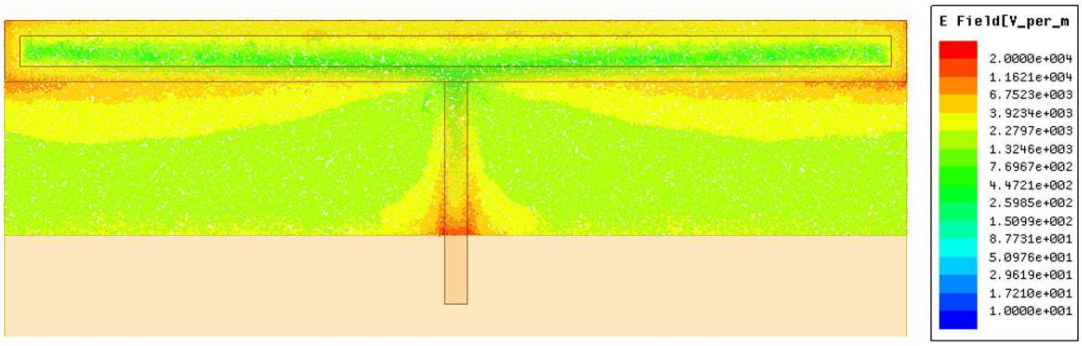

Figure 84 E-Field Distribution at 750MHz for Hollow Printed CE

The MNs for both LB and HB antenna ports were designed with Optenni Lab software (Figure 85). For the optimization of the LB antenna MN, it was intended to have a low reflection coefficient in the LB and a band-stop behavior in the HB. The same goals have been evaluated for the HB, having a stop-band in the LB. The component values were then fine-tuned with the full-wave electromagnetic simulator taking into account the linking transmission lines and vias.

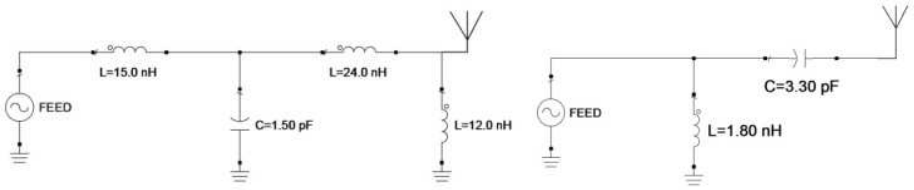

Figure 85 MN Topologies for the Printed CE Antenna (LB on left, HB on right)

The simulated input impedance of the printed CE alone can be seen in Figure 86 with and without the MN. The simulated S-parameters for the whole structure including the HB antenna can be observed in Figure 87. The target frequency bands are covered with a reflection coefficient below -6dB except the frequency interval between 700 and 720MHz. The achieved isolation level is higher than 23dB through both bands which is a consequence of both the antenna placement and the band-stop behavior of the two MN's.

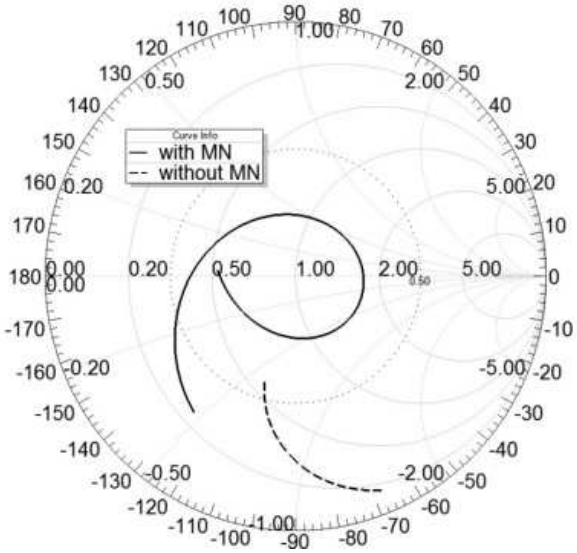

Figure 86 Simulated Input Impedance of the Printed CE Alone (with and without MN)

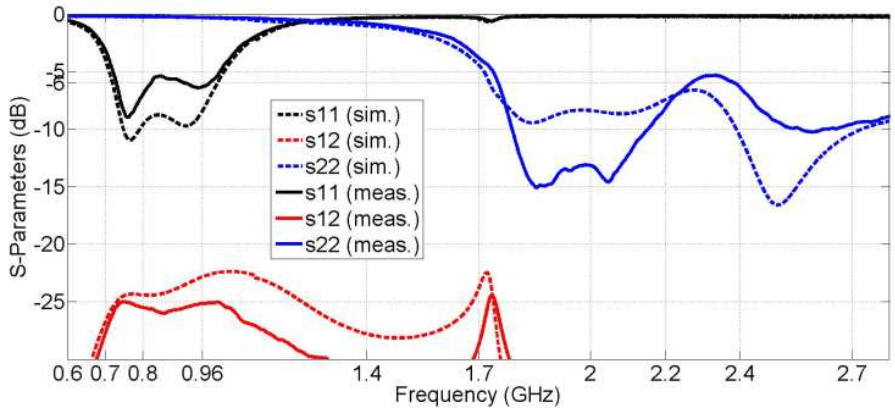

Figure 87 Simulated and Measured S-parameters of the Printed CE Antenna

To explain the wideband behavior of the HB antenna, its input impedance is plotted alone and also when the printed CE is introduced (Figure 88). Additional resonances are created from the coupling between this HB antenna and the LB CE. For instance, the HB antenna is strongly excited at 1.8GHz (Figure 89) whereas at 2.5GHz, the excitation is lower with a stronger coupling with the LB CE. Although there exists some coupling between both antennas at different frequencies, a high isolation is still maintained with the help of the band-stop characteristics of the MNs.

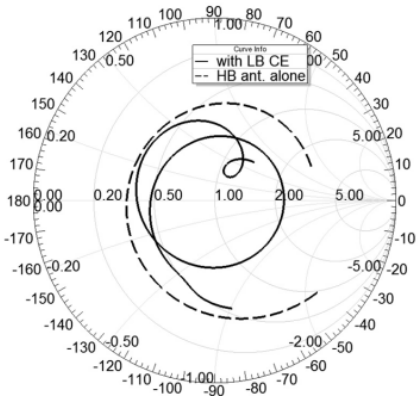

Figure 88 Simulated Input Impedance of the HB antenna alone and with the Printed CE

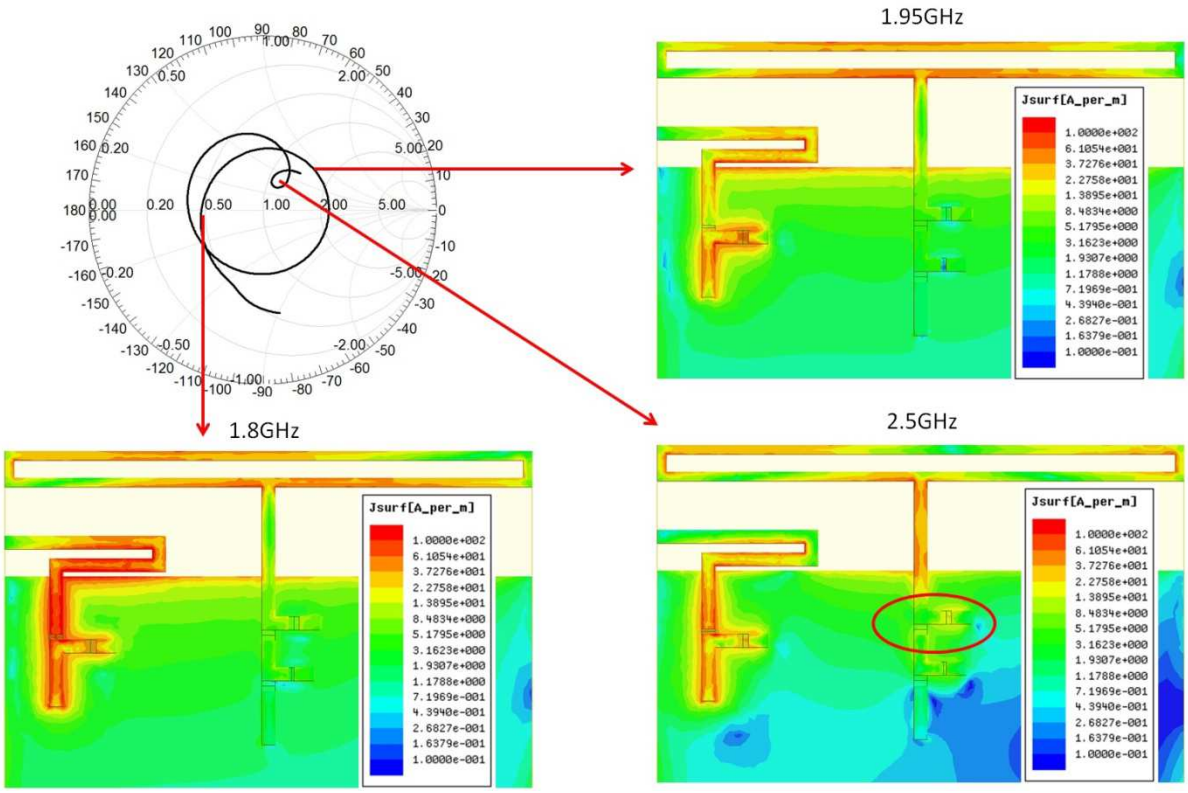

Figure 89 Current Distribution on the Printed CE when the HB Port is Excited

1.3.2. Measurement Results

The printed antenna structure has been manufactured on FR4 substrate as seen in Figure 90. The comparison of the simulated and measured S-parameters can be seen in Figure 87 and on smith chart in Figure 91. There is a fair agreement between the simulated and measured results. The proposed printed antenna can cover the target bands with a measured reflection coefficient below -6dB except the frequency interval of 700-720MHz and also a small overshoot up to -5.5dB around 850MHz.

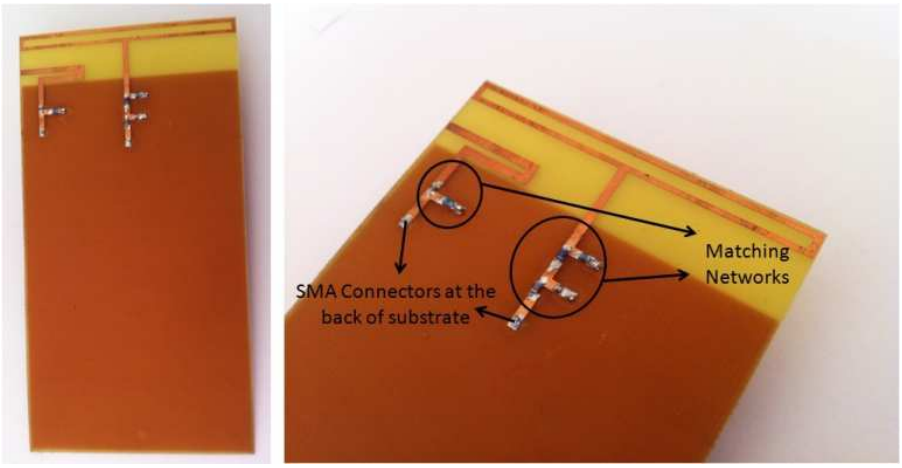

Figure 90 Manufactured Prototype of the Printed CE Antenna

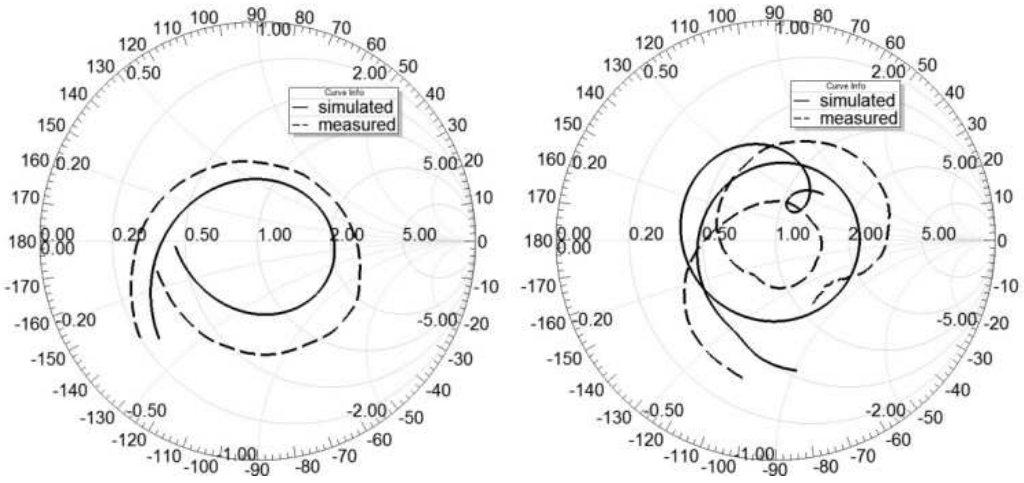

Figure 91 Simulated and Measured Input Impedance of the Printed CE Antenna (LB on left, HB on right)

The total efficiency was measured to vary from -5dB to -1.5dB in the LB and from -4dB to -1.2dB in the HB. The simulated and measured 3D gain radiation patterns are also in good agreement as seen in Figure 93.

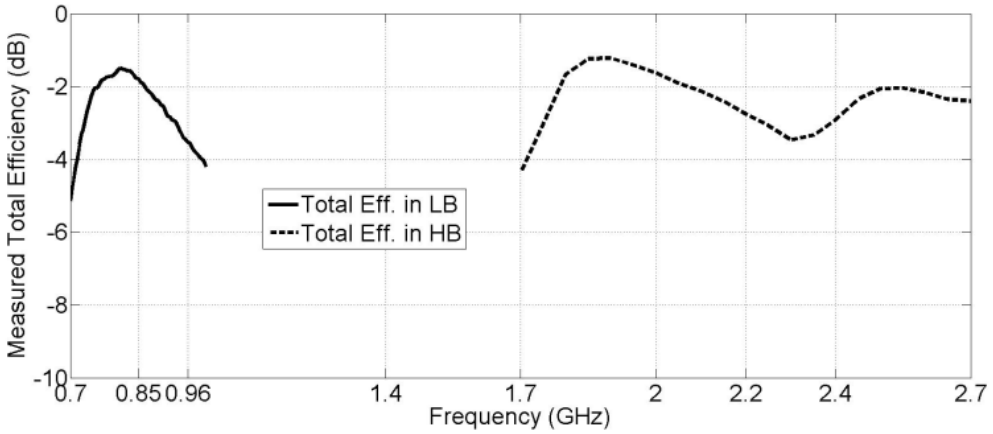

Figure 92 Measured Total Efficiency of the Printed CE Antenna

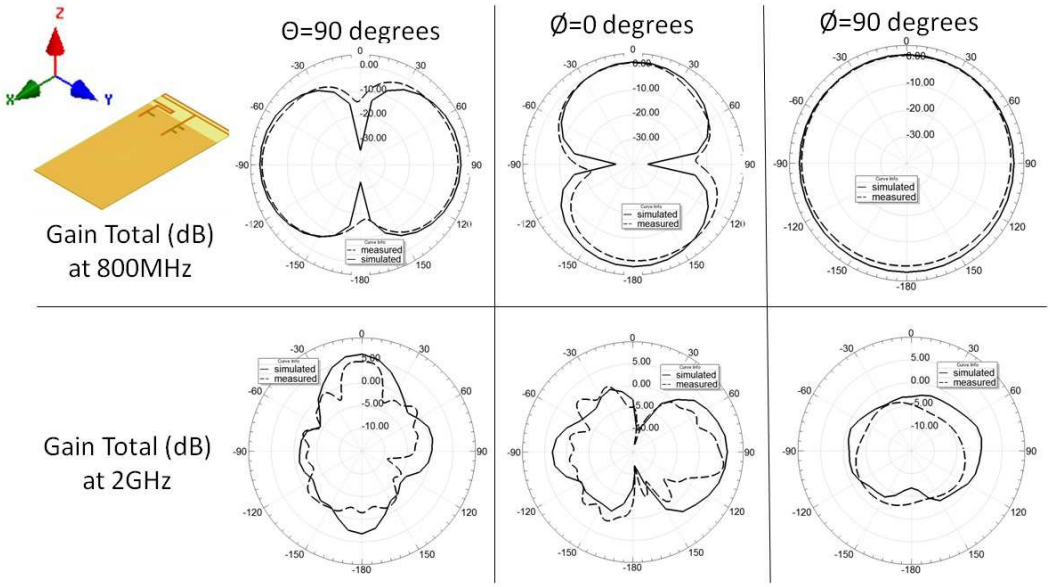

Figure 93 Simulated and Measured Total Gain Patterns for Different Cut Planes

2. TWO-ELEMENT MIMO ANTENNA USING HOLLOW COUPLING ELEMENTS

The newly proposed hollow CE was evaluated in a dual-MIMO antenna scenario as an application example. The target frequency band is 700-960MHz (for a reflection coefficient below -6dB) for both antenna structures. It is aimed to have an isolation better than 10dB between the feeds and an ECC (envelope correlation coefficient) lower than 0.5. Similar studies have been done like in [3-5] to achieve MIMO operation in mobile terminals. In [3], two meandered monopoles were placed on the short edge of a mobile terminal PCB with a hybrid coupler at the feeds to improve isolation. With this antenna, good matching bandwidth and good isolation was achieved only in a very limited bandwidth (between 705-725MHz). A similar structure having two meandered monopoles is also presented in [4], using a suspended line to increase isolation. This antenna is also narrowband, with MIMO operation only in LTE band-13 (746-787MHz). In [5], first an active antenna structure is proposed to cover 700-960MHz with four different states and then the MIMO performance is investigated with combinations of active-passive antennas.

As widely known, it is a challenging problem to obtain a MIMO antenna with low ECC (envelope correlation coefficient) and high isolation for a typical mobile terminal in whole LB (700-960MHz), where the antenna is electrically small and most of the radiation comes from the excitation of the ground plane. Due to this fact, it is hard to obtain a high level of isolation between the two antenna feeds. Although a good isolation in terms of S_{21} is obtained, it is not sufficient to get an acceptable Envelope Correlation Coefficient (ECC) since the radiation patterns should also be different for the two antennas. For this purpose, two identical hollow CEs were placed on the two corners of the short edge of the PCB as in [6]. In this way, a good ECC from the 3D pattern point of view can be expected. To increase the port-to-port isolation between the two CEs, the neutralization technique, which is explained in [7] was used.

2.1. THE LTE MIMO ANTENNA MODEL

The simulation model of the antenna can be observed in Figure 94. An FR4 substrate of 0.8mm thickness was used as the PCB. The height of the antenna is 4.2mm above the substrate. The CEs on each corner are identical, each with a MN of 4 SMD components at the antenna feeds (Figure 95). To increase the inductive effect of the neutralization line, an inductor of 100nH was used in the middle of this line.

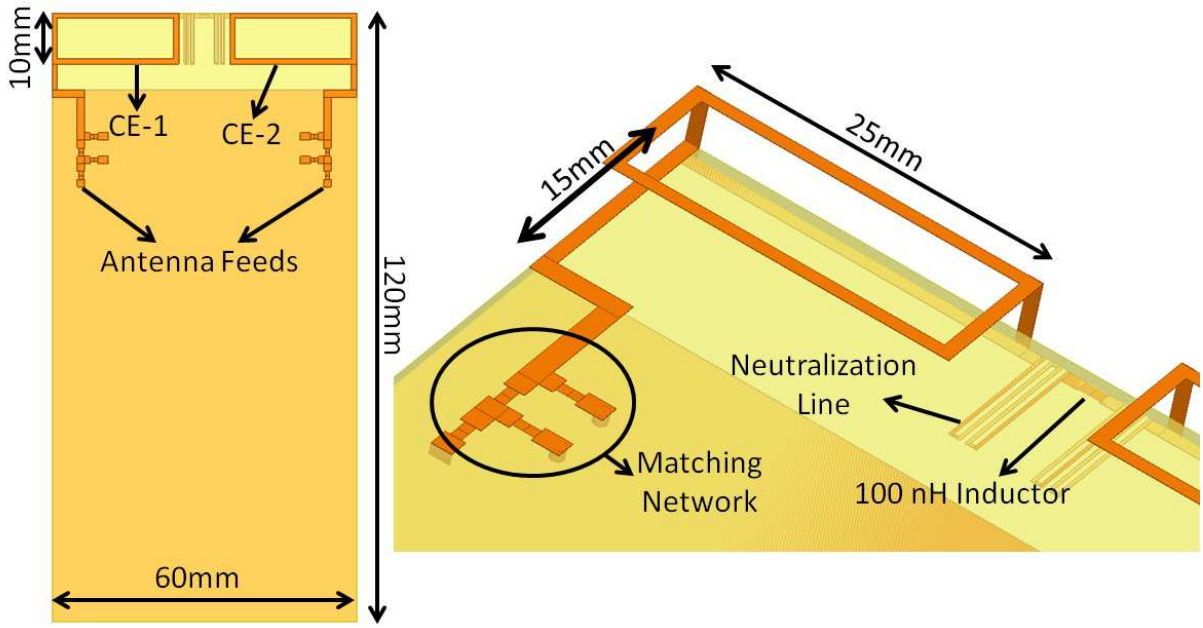

Figure 94 The LTE MIMO Antenna Simulation Model

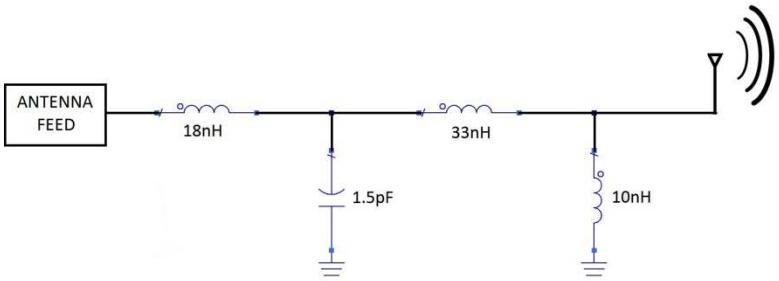

Figure 95 The MN for LTE MIMO CEs

To illustrate the effect of the MN on the input impedance, simulation results have been plotted in Figure 96, between 700-960MHz. As shown with the red curve, the single CE is far from being matched without the MN. The MN (optimized using Optenni Lab software), creates an optimally overcoupled response to obtain the highest possible matching bandwidth.

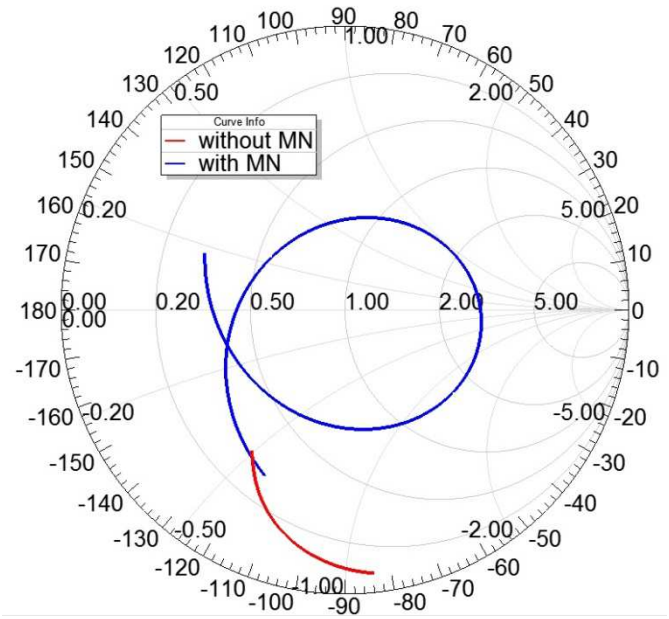

Figure 96 Input Impedance of a Single CE with/without MN

Figure 97 presents the currents on the ground plane when a single CE is placed on top left corner of the PCB. It can be seen that the main direction of current flow on the ground plane is through the long edge of the ground plane, continuing the short edge and having the lowest value close to the corner. The currents flowing on the short edge introduce an offset in the radiation pattern as seen in Figure 98 (at 800MHz). The pattern is different from regular patterns in the sense that it is slightly rotated through the diagonal direction of the PCB, which is an interesting feature to obtain a reasonable ECC value when the second CE is added.

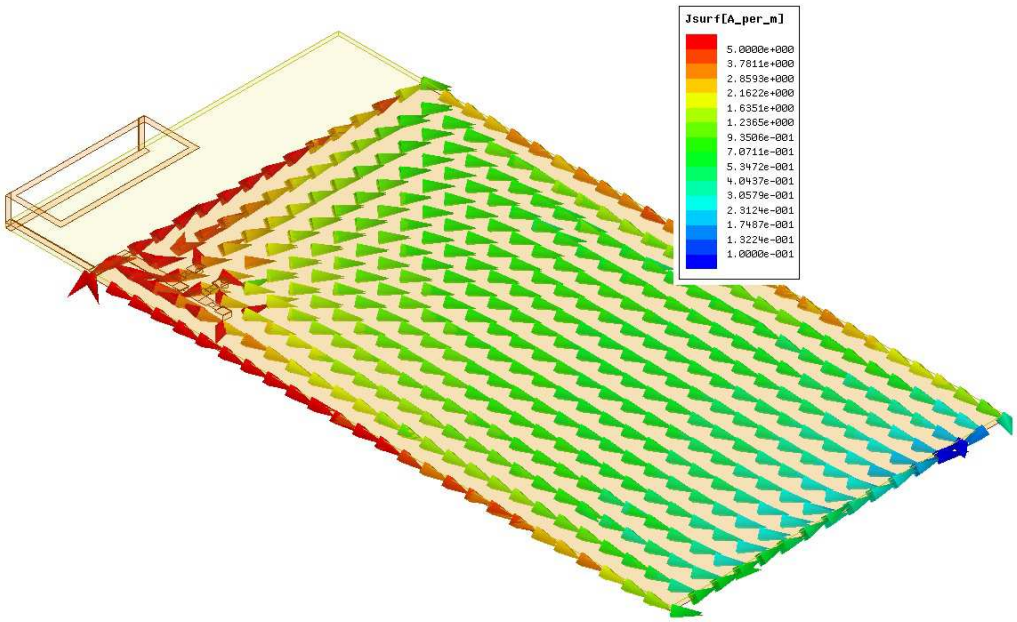

Figure 97 Current Distribution on the Ground Plane of a Single CE with MN

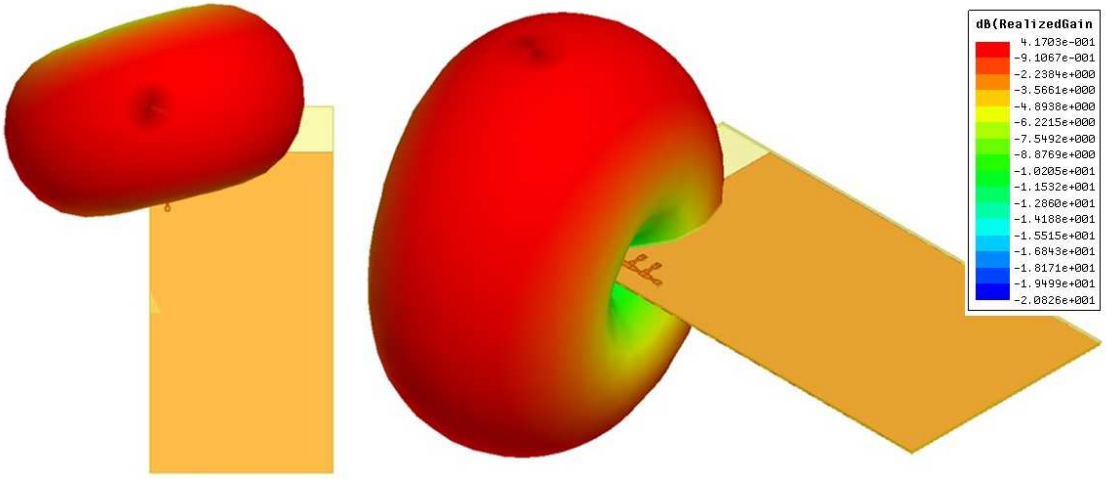

Figure 98 3D Radiation Pattern of a Single CE placed at the Top Left Side of the PCB

Another very important point for this antenna is the neutralization line added between the two CEs, connecting them inductively to increase the isolation between their feeds. The main idea behind this technique is to create an additional inductive path (to the existing capacitive coupling path). In this way, the currents between the two antennas are forced to flow somewhere else than the non-excited feed. Detailed description of the methodology is not given here for the sake of brevity. The antenna (consisting of both CEs on each corner with MNs) was simulated both with and without the neutralization line and the results are presented in Figure 99. The isolation is poor in the absence of the neutralization, falling down to 4.5dB around 720MHz. After the neutralization line is introduced, a significant improvement is seen in the isolation, being better than 10.2dB through the target band (30% bandwidth). The simulated reflection coefficient for the CEs (with MNs) is below -6dB between 740-940MHz.

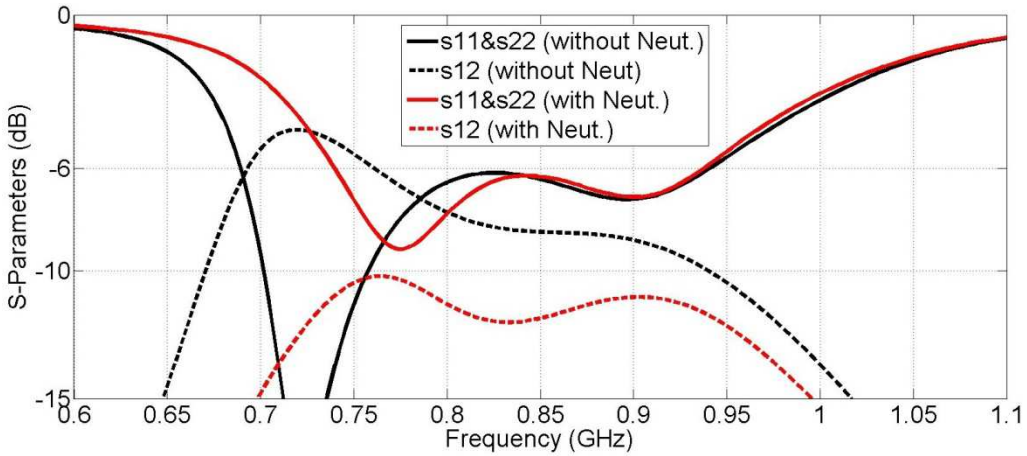

Figure 99 Effect of the Neutralization Line on Simulated S-Parameters

The simulated 3D radiation patterns for the whole antenna system (at 800MHz) can be observed in Figure 100, where the pattern for CE-1 is plotted when the other port is matched and vice versa. The rotated dipole type patterns, symmetrical along the longitudinal axis, for the two CEs verify the particular excitation of the ground plane. It produces pattern diversity which is of particular interest for low ECC.

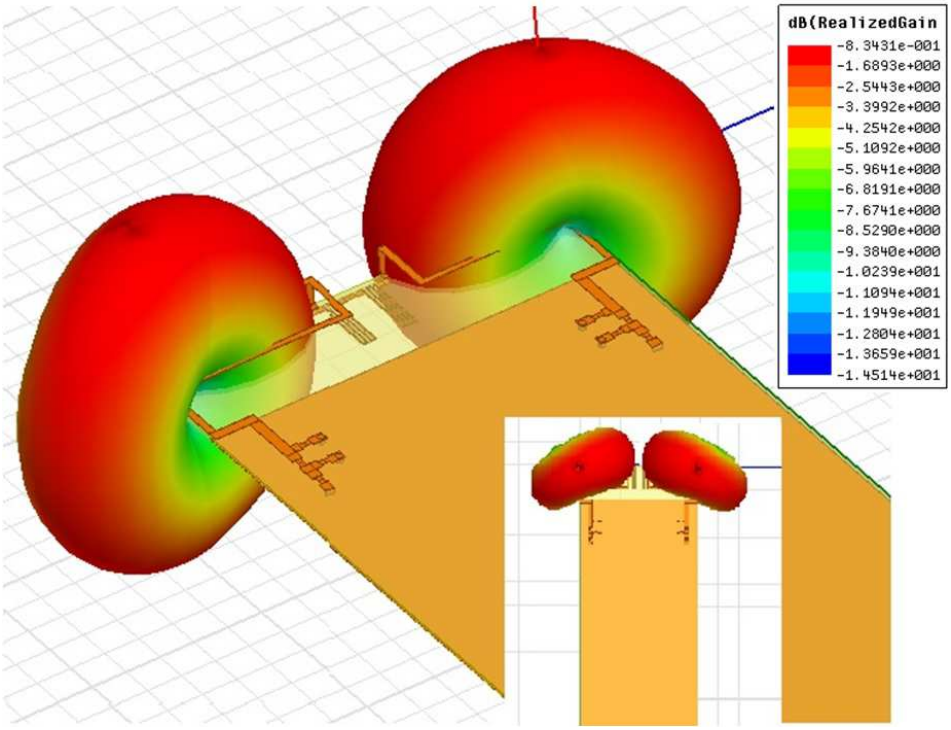

Figure 100 3D Radiation Pattern of Dual CE Antenna

2.2. MEASUREMENT RESULTS OF THE LTE MIMO ANTENNA

The proposed antenna has been manufactured on FR4 substrate (Figure 101). The S-parameter measurements have been performed, whose results are presented in Figure 102. A fair agreement can be observed between simulation and measurements. The frequency region between 740-950MHz can be covered with a reflection coefficient below -6dB and the isolation between the two ports is always higher than 11dB in the measurements. From the best knowledge of the authors, this isolation level was not achieved over such a broad band for LTE mobile phones.

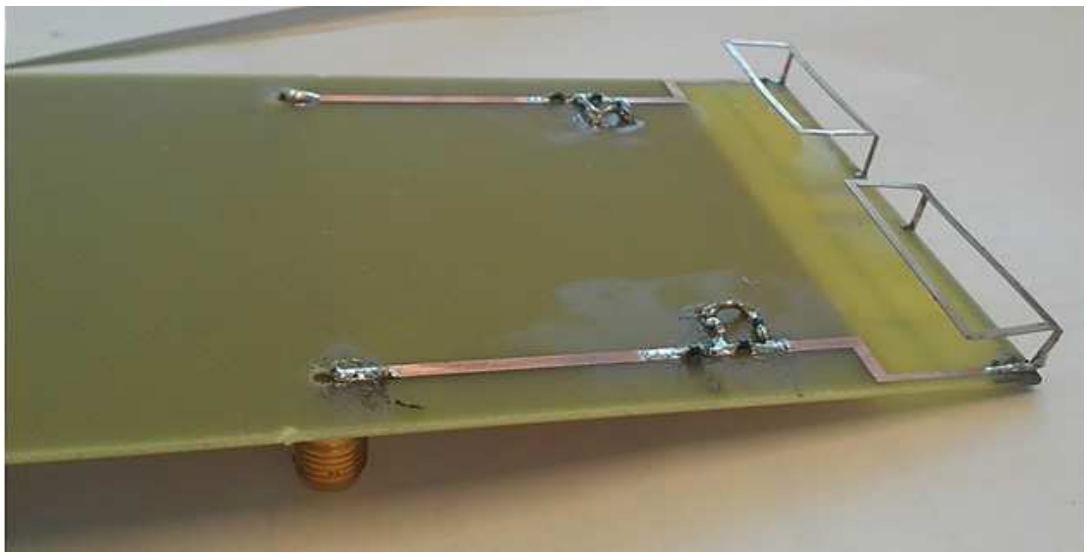

Figure 101 Manufactured MIMO Antenna Prototype

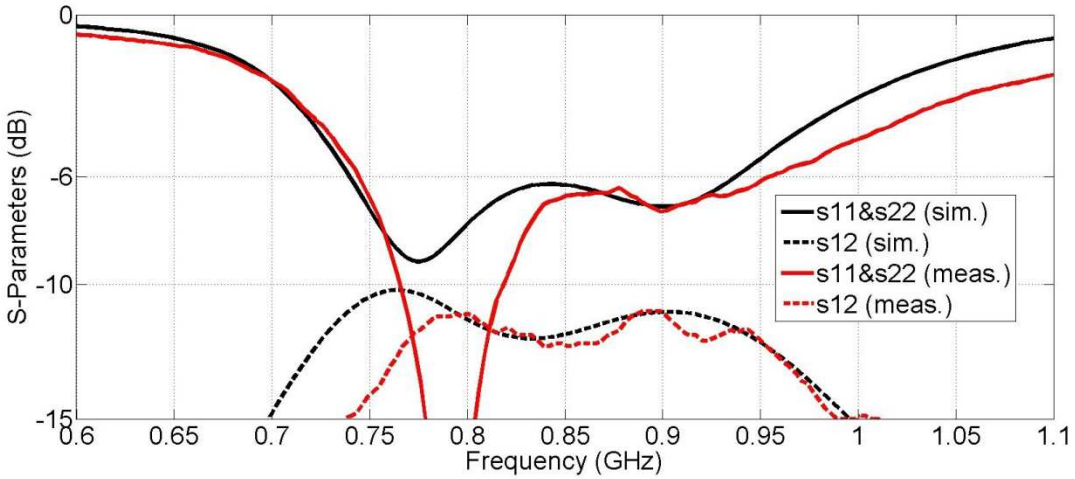

Figure 102 Simulated and Measured S-Parameters of the MIMO Antenna

The total efficiency of the MIMO antenna system was measured in a Satimo Starlab station, where the unmeasured antenna port is terminated with a matched load. The two CEs give similar performance in terms of efficiency so the result of only one CE is presented in Figure 103. The antenna has a measured total efficiency between -6dB and -4dB in the coverage band.

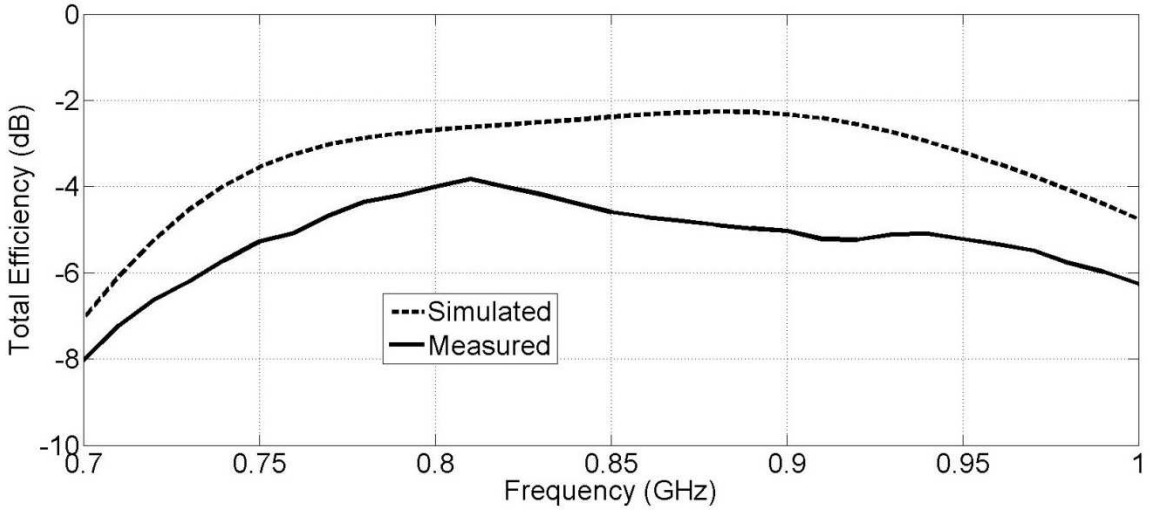

Figure 103 Simulated and Measured Total Efficiency of the MIMO Antenna

As widely known, there are two formulas used for the calculation of the envelope correlation coefficient of MIMO antenna systems. The first one presented in equation-1 uses the S-parameters of the two antennas whereas the second one (equation-2) uses the 3D complex radiation pattern. Although much simpler to calculate, Equation-1 is not suitable in this study since it may give misleading results due to the lossy antennas. For this reason, equation-2 (for an isotropic propagation environment) was used to calculate the ECC.

$$\rho_e = \frac{|s_{11}^{*}.s_{12}+s_{21}^{*}.s_{22}|^2}{(1-|s_{11}|^2-|s_{21}|^2)(1-|s_{22}|^2-|s_{12}|^2)} \qquad \text{Equation-1}$$

$$\rho_e = \frac{\left|\iint_{4\pi} F_1(\theta,\phi)^*.F_2(\theta,\phi)d\Omega\right|^2}{\left(\iint_{4\pi}|F_1(\theta,\phi)|^2 d\Omega\right).\left(\iint_{4\pi}|F_2(\theta,\phi)|^2 d\Omega\right)}$$ Equation-2

The comparison of the simulated and measured ECC using the radiation patterns is given in Figure 104. The agreement between simulation and measurement (complex radiation patterns have been recorded with the SATIMO station) seems very good. The measured ECC turned out to be below 0.5 (a generally accepted value for mobile terminals) between 765MHz and 950MHz which contains most of the coverage band. The MIMO antenna system is suitable for MIMO applications in mobile terminals, keeping in mind that other metrics like diversity gain and mean effective gain should also be measured.

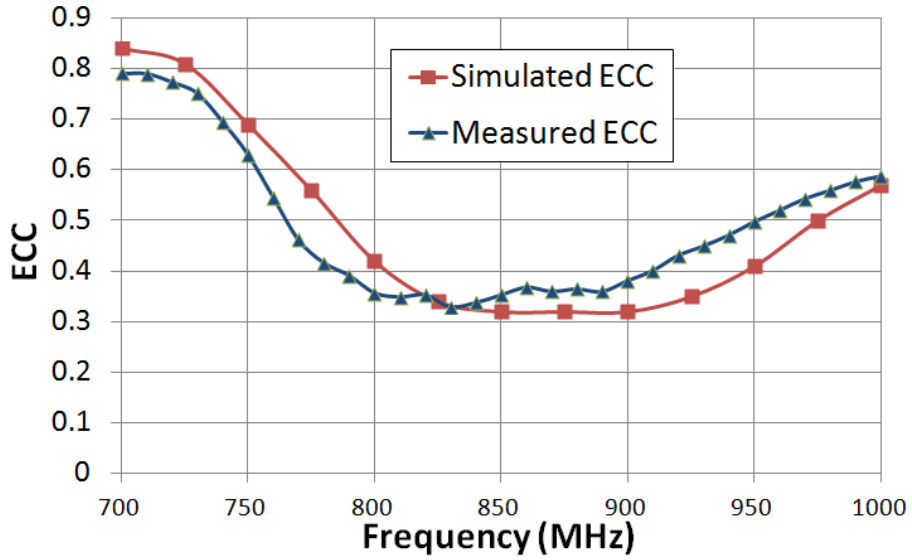

Figure 104 Simulated and Measured ECC of the MIMO Antenna

3. CONCLUSION

A novel approach using CEs for 700-960MHz and 1.7-2.7GHz coverage in mobile terminals was presented in this chapter. The approach consists of making the plain CE hollow by removing the metal plate inside and leaving a metal strip of around 1mm. In this way, a space efficient layout is obtained with weaker E-fields inside the hollow region. This space is then used for inserting a HB antenna separately fed with a high isolation between the feeds. A comparison between plain CE and hollow CE was first presented showing the negligible difference in terms of bandwidth potential performance.

The approach was then evaluated with a CE of 3mm height to cover the 700-960MHz frequency band. A monopole antenna was inserted inside this hollow CE on the substrate and it was shown that the whole target band can be covered with this dual feed structure. Another antenna was proposed in section 1.3, using again a dual-feed topology but this time the CE was also printed directly on the substrate.

Finally, in section 2, the CE idea was applied to obtain MIMO performance in the LB, using two CEs situated on the two corners of the PCB. With this layout, specific currents of the ground plane were excited for both CEs, enabling a level of default isolation. To further improve the isolation, the

neutralization line technique was applied to the CEs. The S-parameter and ECC measurements were performed and presented for the MIMO antenna.

It was shown in this chapter that it is possible to obtain the coverage of 4G communication standards in a mobile terminal with coupling element antennas. The disadvantage of the prototypes presented here can be the 3D structure without any supporting materials. This might lead to some manufacturing errors in different prototypes of the same antenna, due to tolerances. For this reason, antenna structures with a different manufacturing technology will be addressed in the next chapter, which will be capable of supporting 3D shapes for antennas.

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CHAPTER IV
SINGLE FEED ANTENNAS ON LDS TECHNOLOGY

This chapter presents two antenna designs which are realized on plastic material using the Laser Direct Structuring (LDS) technology. Both of these antennas utilize the parasitic strip approach (presented in Chapter II) for bandwidth enhancement. The first antenna to be presented covers the low-LTE and UMTS frequency bands of 700-960MHz and 1.7-2.2GHz, while the second antenna can cover the same bands also including the WLAN2400 and LTE2500 bands (i.e. 700-960MHz and 1.7-2.7GHz in total).

A novel point here is the realization of the antennas using Laser Direct Structuring (LDS) technology which offers more freedom in terms of antenna layout and also enables the utilisation of the available space more efficiently. As also described in [1], this process uses a thermoplastic material which has already been doped with a metal-plastic additive(Figure 105-a). A laser beam then tracks the surface of the material, according to the shape of the traces intended on it(Figure 105-b). In this way, a microscopically rough surface is created, forming the nuclei for the metallization step(Figure 105-c). Finally, the metal traces are formed by putting the laser activated material in the electroless copper baths(Figure 105-d). Using this technology, a high degree of freedom is obtained in terms of the metallization shape.

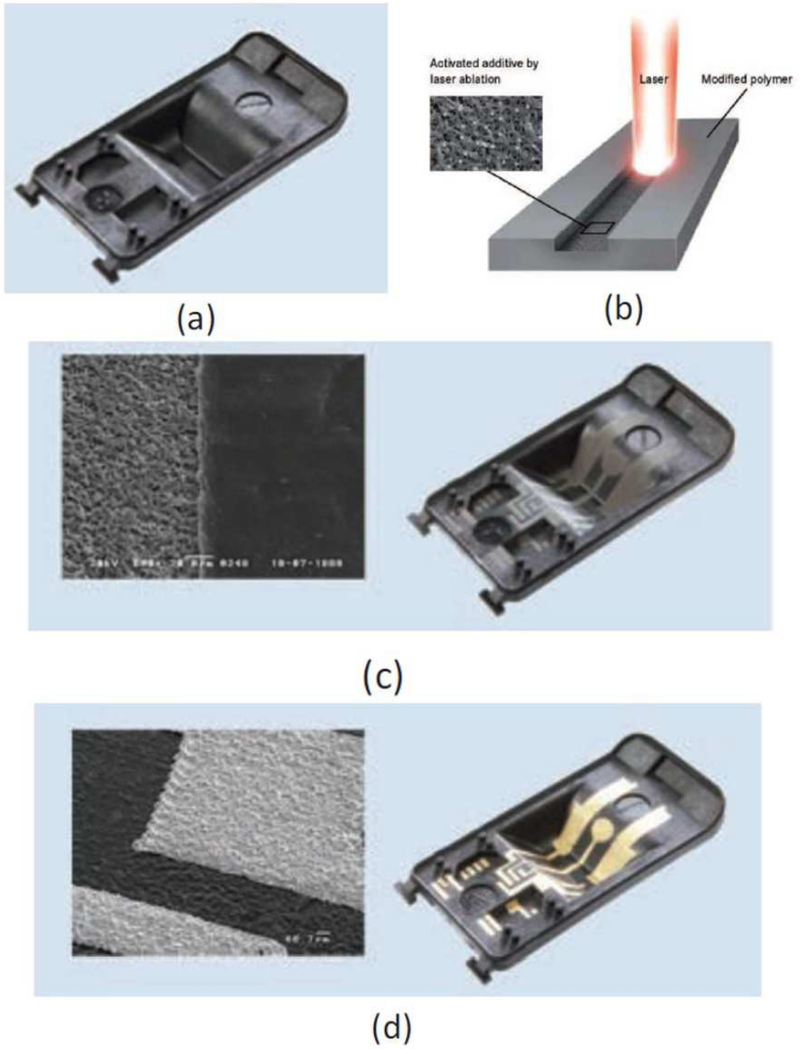

Figure 105 LDS Manufacturing Process

As also mentioned in the Introduction part, many antennas using parasitic elements for LTE/4G frequency bands have been proposed[2-5]. The common property of these antennas is their

relatively complex geometry, which might make the antenna performance sensitive to manufacturing tolerances. Here, the main target is to simplify the geometry of the antenna as much as possible by introducing a matching network having few number (targeted as less than 4) of SMD components.

1. LOW-LTE AND UMTS ANTENNA

1.1. ANTENNA MODEL

The antenna model can be observed in Figure 106. For the plastic piece, DP-T-7140 material [6] was chosen since it is appropriate for reflow soldering, considering the possible usage of lumped components on the plastic piece. This plastic piece is shown in grey colour in Figure 106 and it has a relative permittivity of 4.1 and a loss tangent of 0.0138, given by the manufacturer for 1MHz. The dimensions of the plastic mold are 57mm X 18mm with a height of 5mm. The plastic mold rests on a PCB of FR4 substrate with the dimensions 114mm X 59mm X 0.8mm. There is a ground plane of 75mm X 50mm under the PCB. The main reason of choosing 75mm for the length of the ground plane is to simulate a worst case scenario, where the antenna relies only on the 3.5 inch touch screen of a smart phone, as the ground plane. This is a generic scenario for today's smart phones so the antennas designed are able to be integrated in any smart phone. One point to consider is the additional components (which will have ground connection) in the vicinity of the touch screen so the total ground plane length might be subject to an increase. This increase in the ground plane length will shift its resonance frequency down, enabling easier coverage of the LB. For example, increasing the length of the ground plane from 75mm to 100mm will shift the fundamental wavemode frequency from 1.45GHz to 1.18GHz.

The connection between the metallizations on the plastic mold and the traces on the PCB are achieved by using two pogo pins having 3mm height, under the plastic mold. The antenna basically consists of a driven strip (shown in orange), which capacitively excites a parasitic strip (shown in red) connected to ground on one end. The design methodology when optimizing the final form of the antenna was to tune the length of the parasitic strip, in combination with the driven strip, until obtaining sufficient bandwidth potential for dual-band operation. A meander was introduced to the parasitic element for this reason in the final antenna form. Concerning the input impedance of the antenna, a matching network consisting of four lumped components was decided to be used at the antenna feed.

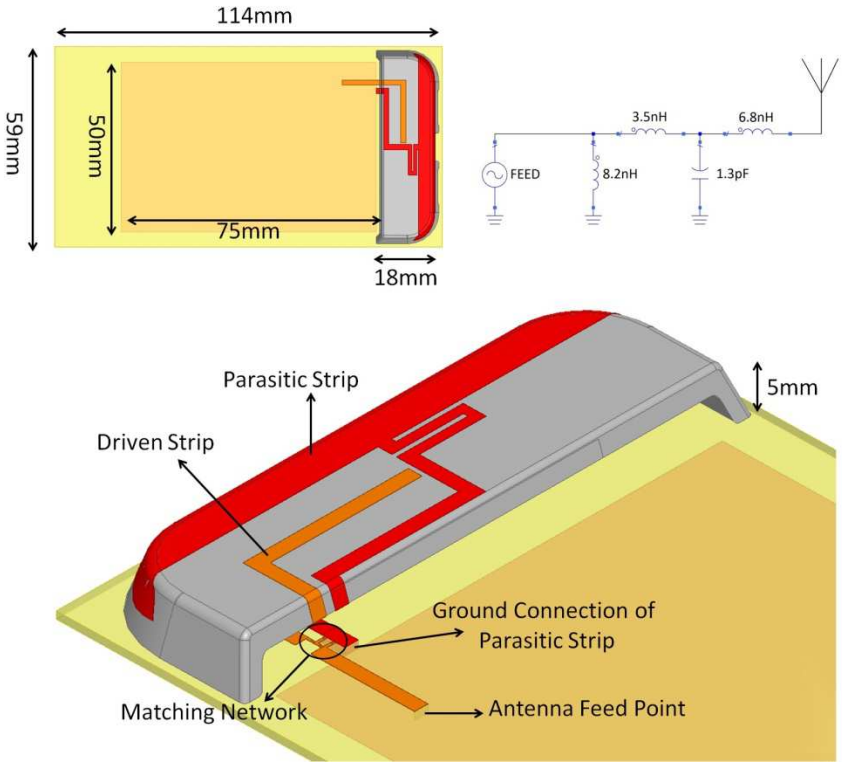

Figure 106 UMTS Antenna Model

To show the effect of the parasitic element on the input impedance of the antenna, simulations were performed in the case of "driven strip only" and "whole structure without MN". These simulation results can be observed in Figure 107, where the LB impedance is plotted between 700-960MHz and HB response plotted between 1.7-2.2GHz. It can be seen from the figure that introducing the parasitic strip creates a resonance in the LB, increasing the real part of the input impedance considerably. This behaviour enables an increase in the bandwidth potential of the antenna especially in the LB (Figure 108).

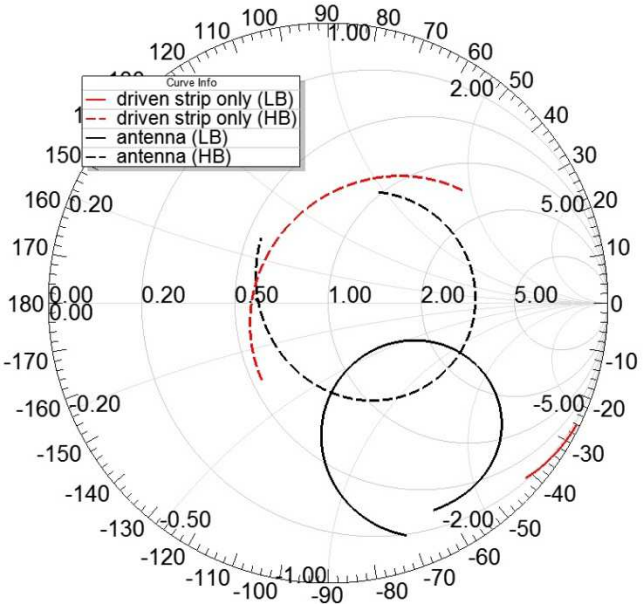

Figure 107 Simulated Input Impedance of the UMTS Antenna, with/without Parasitic Strip

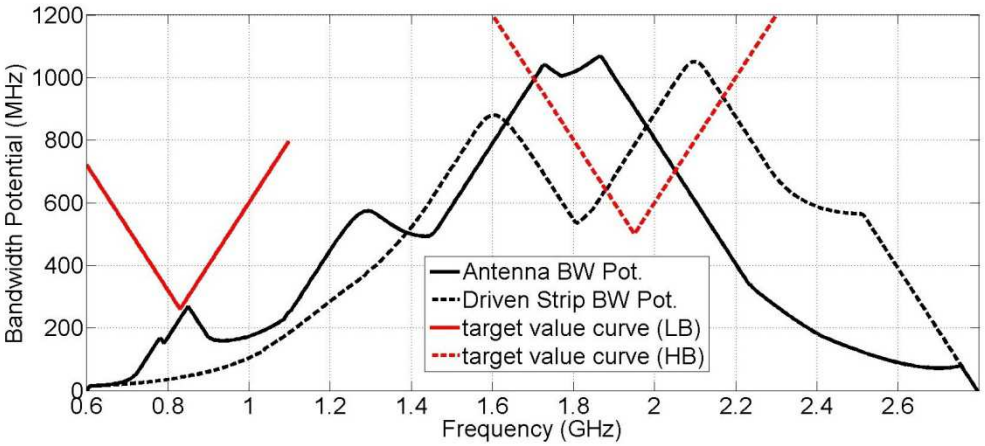

Figure 108 Bandwidth Potential of the UMTS Antenna, with/without Parasitic Strip

The effect of the parasitic element can be also seen looking at the surface current distribution in Figure 109. At 750MHz, a strong excitation of the parasitic element is observed, while its excitation is weaker at 900MHz. It can also be seen that, at both frequencies, the currents on the ground plane are excited almost at the same strength, which enables the exploitation of the low-Q property of the thick ground plane.

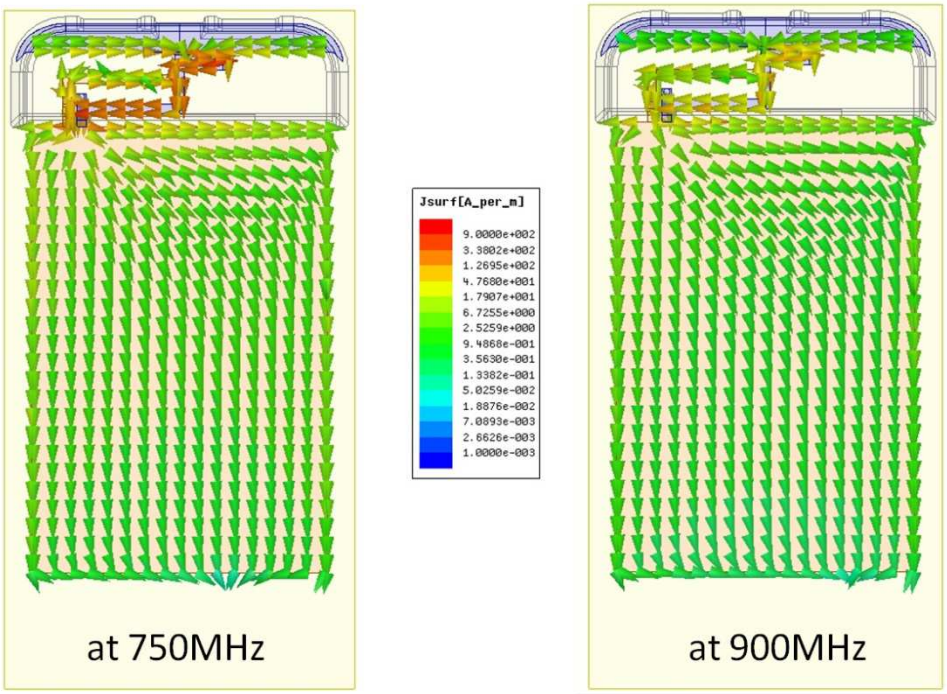

Figure 109 Current Distributions on the UMTS Antenna and Ground Plane

The simulated antenna impedance with and without MN can be observed on smith chart in Figure 110. The corresponding reflection coefficient can also be observed in Figure 111. The antenna can cover the target bands of 700-960MHz and 1.7-2.2GHz, although a small overshoot in reflection coefficient is observed around 780MHz rising up to -5dB (in simulations).

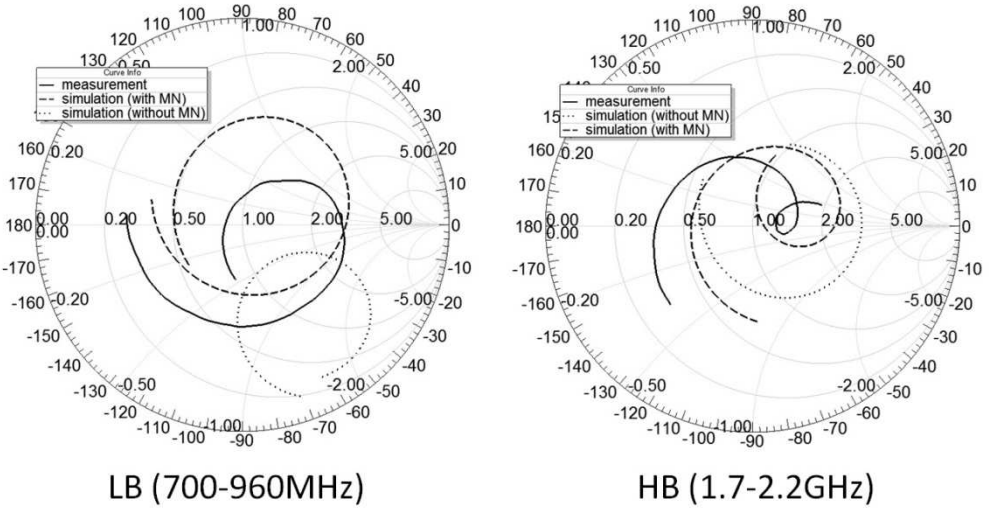

Figure 110 Simulated and Measured Input Impedance for UMTS Antenna, with/without MN

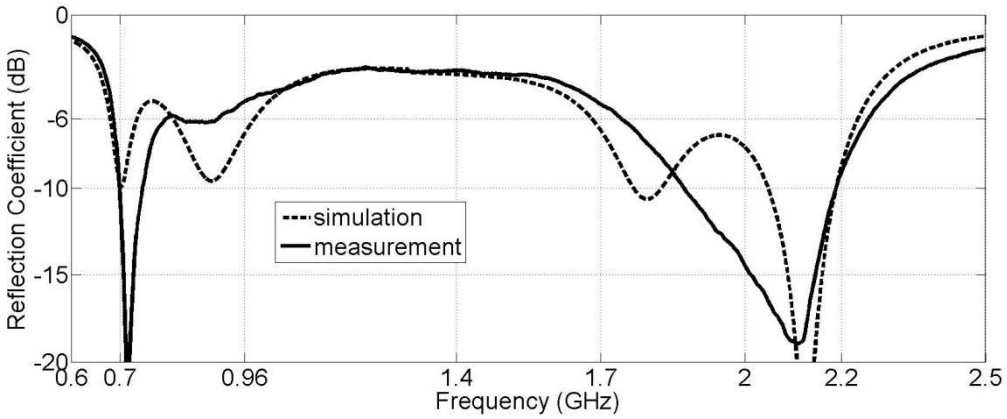

Figure 111 Simulated and Measured Reflection Coefficient for the UMTS Antenna

1.2. MEASUREMENT RESULTS

The manufactured antenna (using LDS technology) can be seen in Figure 112. For the antenna feed, an SMA connector was placed on the PCB as shown in left in Figure 112. To validate the simulation results, s-parameter measurements were performed with the manufactured prototype and are presented in Figure 110 and Figure 111. A fair agreement can be observed between simulated and measured reflection coefficients. The antenna has a measured reflection coefficient below -6dB in the LB and HB with small overshoots around 950MHz and 1.7GHz.

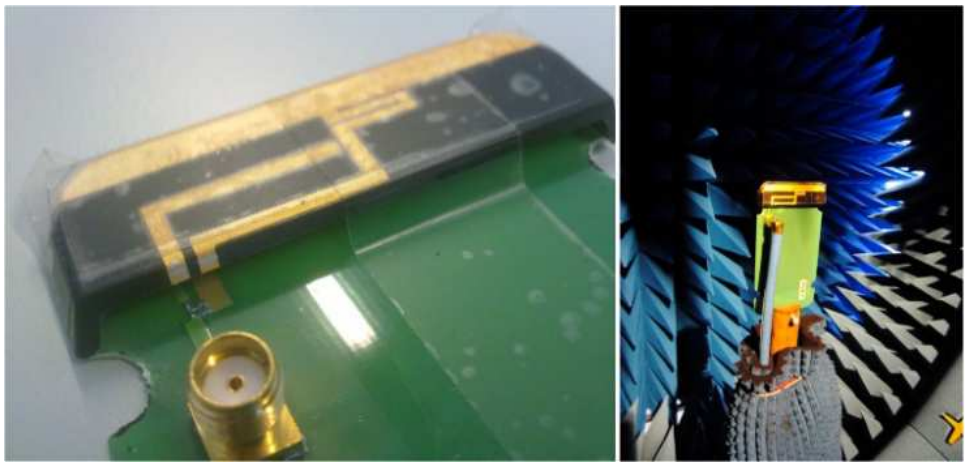

Figure 112 Manufactured UMTS Antenna Prototype and Efficiency Measurement Environment

The total efficiency measurements were also performed in a Satimo Starlab station. Two configurations were measured with this prototype, consisting of short (75mm) and long (90mm) ground plane dimensions. For the long ground plane configuration, the ground plane was extended to 90mm from 75mm by using copper tapes under the PCB. The comparison of the measured reflection coefficient for short and long GND plane configurations is seen Figure 113. Extending the ground plane to 90mm helps for slightly better coverage in the LB (around 950MHz) since most of the radiation is coming from the ground plane around this frequency and the chassis resonance is coming more towards the band of interest by increasing the length (1.28GHz instead of 1.46GHz). The effect is not significant in the HB for this antenna & MN combination.

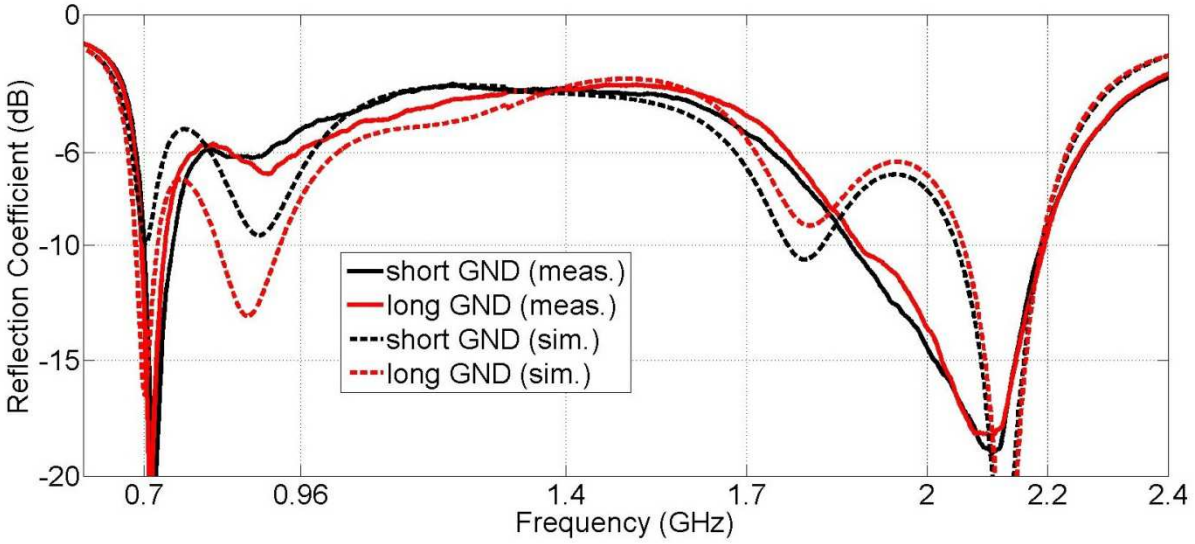

Figure 113 Measured/Simulated Reflection Coefficient of UMTS Antenna with Short/Long Ground Plane

The measured total efficiency for the two configurations can be observed in Figure 114. With the 75mm ground plane, the measured total efficiency is between -7.5dB and -3dB in the LB and between -4dB and -2dB in the HB. Increasing the ground plane length to 90mm has a considerable effect on the total efficiency in the LB, increasing the efficiency by approximately 1.5dB on average. The efficiency in the LB is now between -4dB and -2dB. The efficiency in the HB is not affected by the change, since a major part of the radiation comes from the antenna itself at this frequency range.

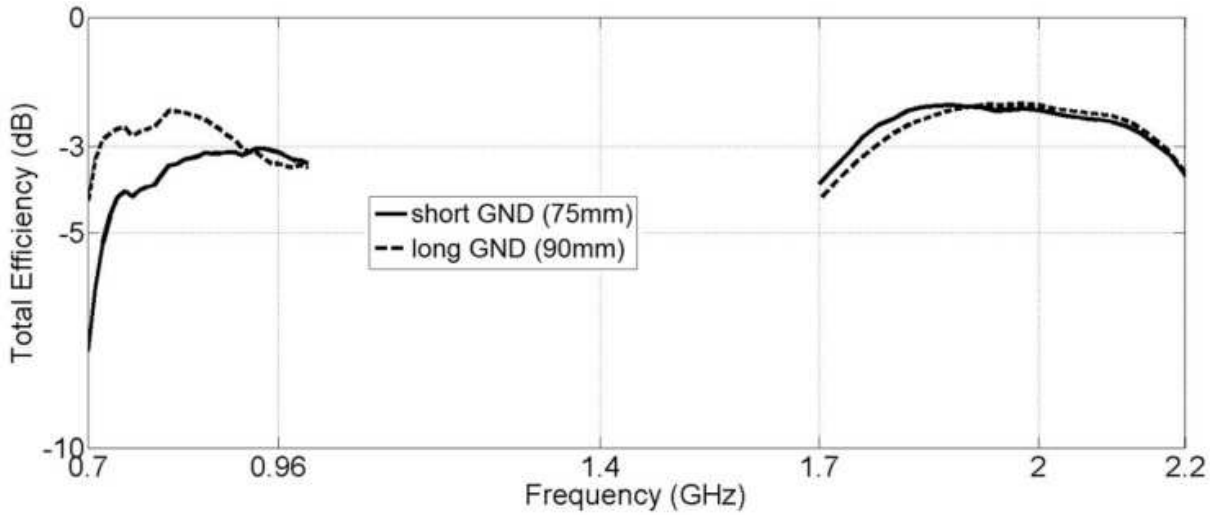

Figure 114 Measured Total Efficiency of UMTS Antenna with Short/Long Ground Plane

2. LTE ANTENNA

2.1. ANTENNA MODEL

A quite similar antenna was designed using the same dimensions for the plastic piece, the ground and the FR4 substrate. The orientation of the driven strip was changed, facing the open end to the corner of the ground plane, to exploit better the low-Q property of the ground plane especially in the HB. The parasitic strip is again excited by this driven strip via capacitive coupling, creating a resonance in the LB, increasing the bandwidth potential. Considering the bandwidth potential of the antenna, a matching network consisting of only two SMD components (a shunt inductor and a series capacitor) was evaluated at the antenna feed but due to the series resonance frequency (SRF) issue, the shunt inductor was split into two inductors, increasing the number of SMD components to three. Pogo-pins were again used under the plastic mold, for the connection of the antenna elements and PCB traces.

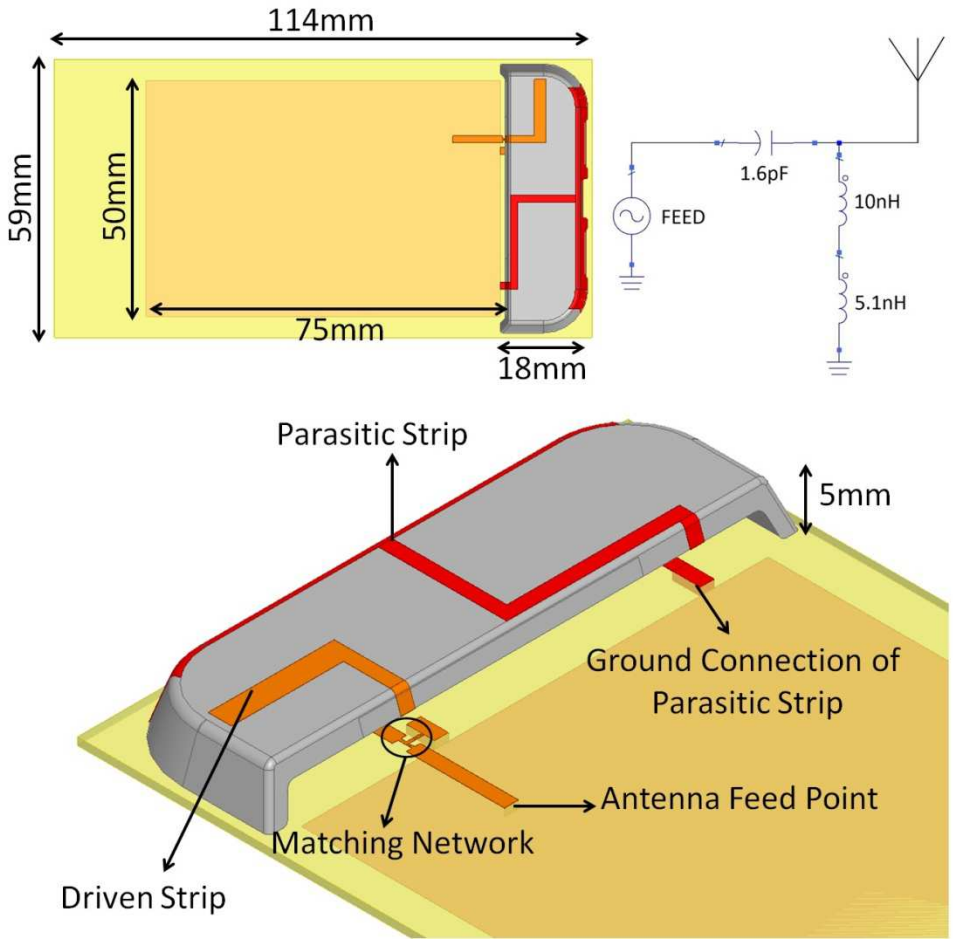

Figure 115 LTE Antenna Model

The simulations were performed for the antenna with and without the parasitic strip (driven strip only). They are presented in smith chart in Figure 116, without any MN at the feed. The introduction of the parasitic element creates a resonance in the LB, increasing the real part of the input impedance. This enables higher bandwidth potential especially in the LB (Figure 117). For the HB response, an additional resonance can also be observed after the introduction of the parasitic element, which occurs approximately around 2GHz (loop in the smith chart).

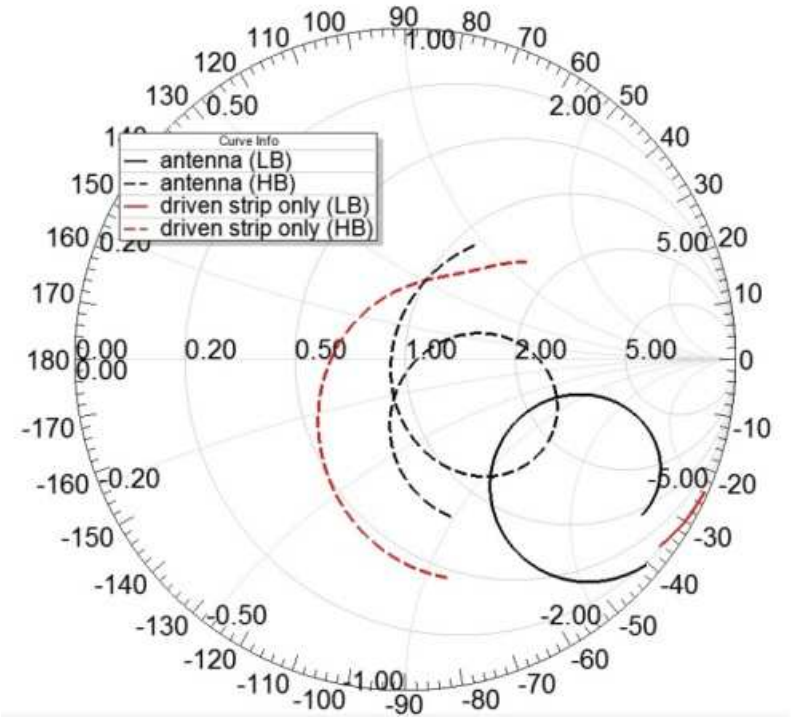

Figure 116 Simulated Input Impedance of the LTE Antenna, with/without Parasitic Strip

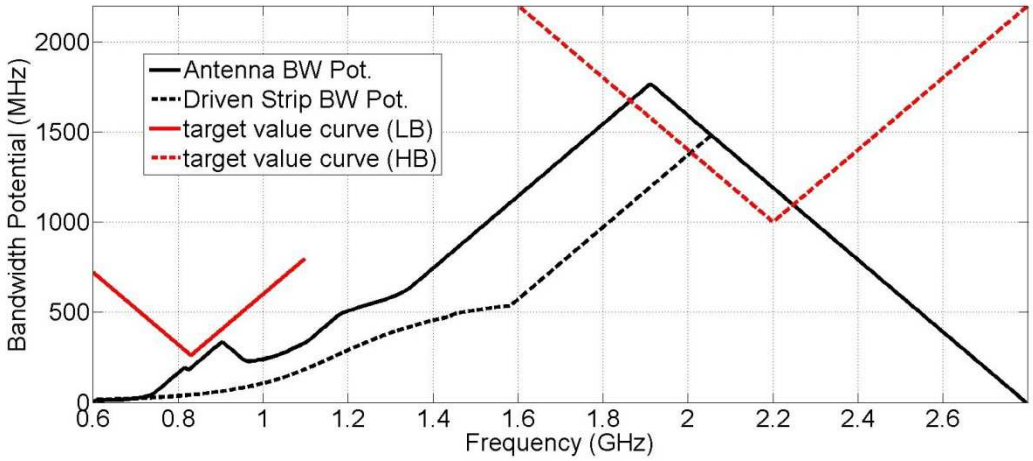

Figure 117 Bandwidth Potential of the LTE Antenna, with/without Parasitic Strip

In order to explain better the role of the parasitic strip in the HB, the input impedance of the antenna structure without MN is given again in Figure 118, with current distributions over the antenna and the ground plane at two different frequency points. There is a loop in the smith chart at 2GHz and when the corresponding surface current is observed, a strong excitation of the parasitic element can be seen. At another frequency point (2.5GHz), the excitation of the parasitic element is weak whereas the driven strip is strongly excited. This suggests that the role of the parasitic element is not confined only within the LB, it also enables higher bandwidth in the HB by creating additional resonance.

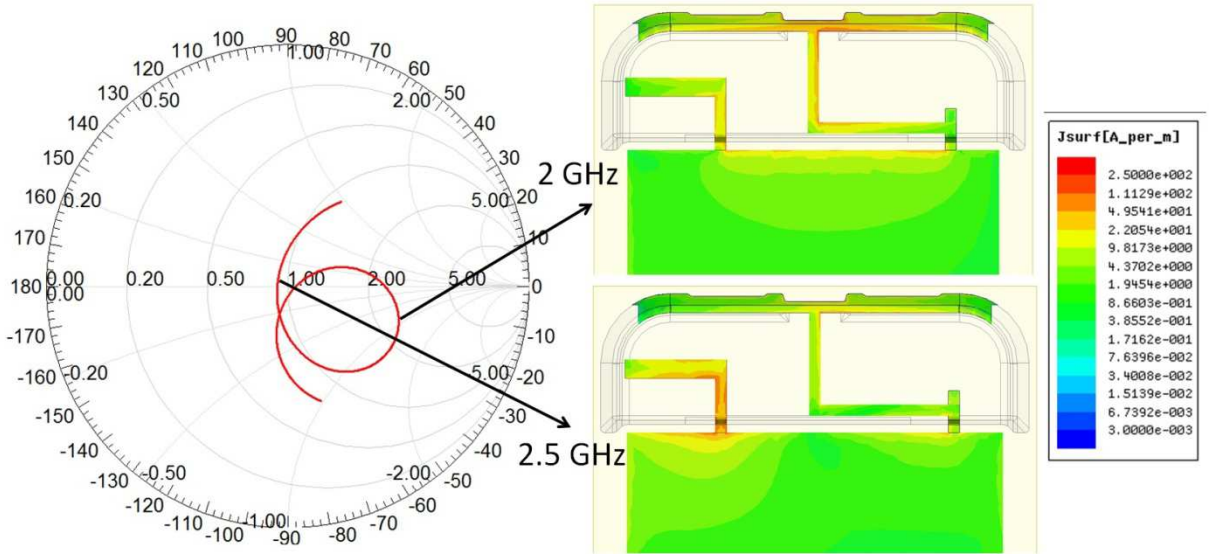

Figure 118 Surface Current Distribution of the LTE Antenna

The simulated input impedance and reflection coefficient of the antenna (with and without MN) are presented in Figure 119 and Figure 120. The antenna can cover the whole target band except some overshoots around 950MHz, 1.7GHz and 2.2GHz.

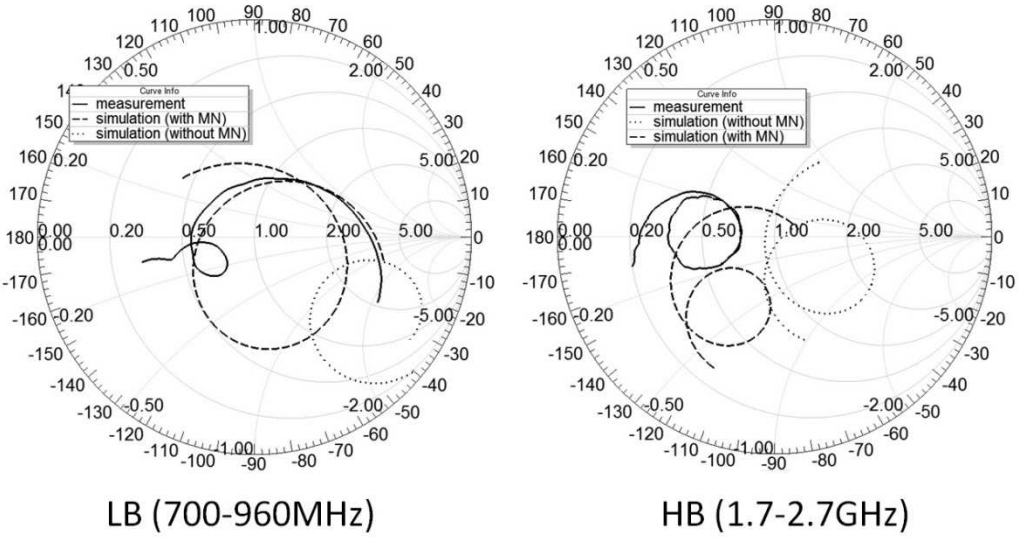

Figure 119 Simulated and Measured Input Impedance for LTE Antenna, with/without MN

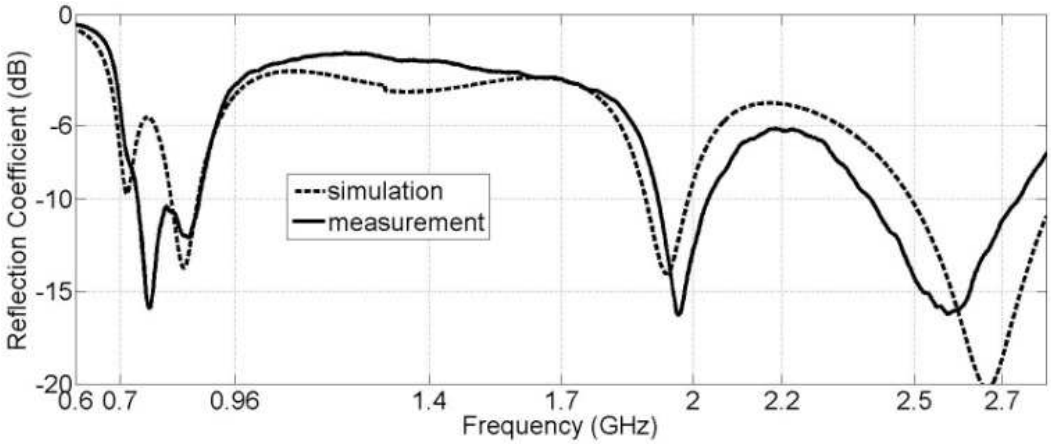

Figure 120 Simulated and Measured Reflection Coefficient of the LTE Antenna

2.2. MEASUREMENT RESULTS

The manufactured antenna prototype can be seen in Figure 121. The antenna is fed through an SMA connector placed on the PCB. The comparison of the measured and simulated reflection coefficient can be seen in Figure 119 and Figure 120. A fair agreement is observed between simulation and measurement results. The agreement is less obvious for LB in smith chart, which may be coming from the misalignment of the plastic piece on the PCB. In the HB, the difference in the input impedance is only a phase difference, coming from the possible differences on the feed line between simulation and measurement.

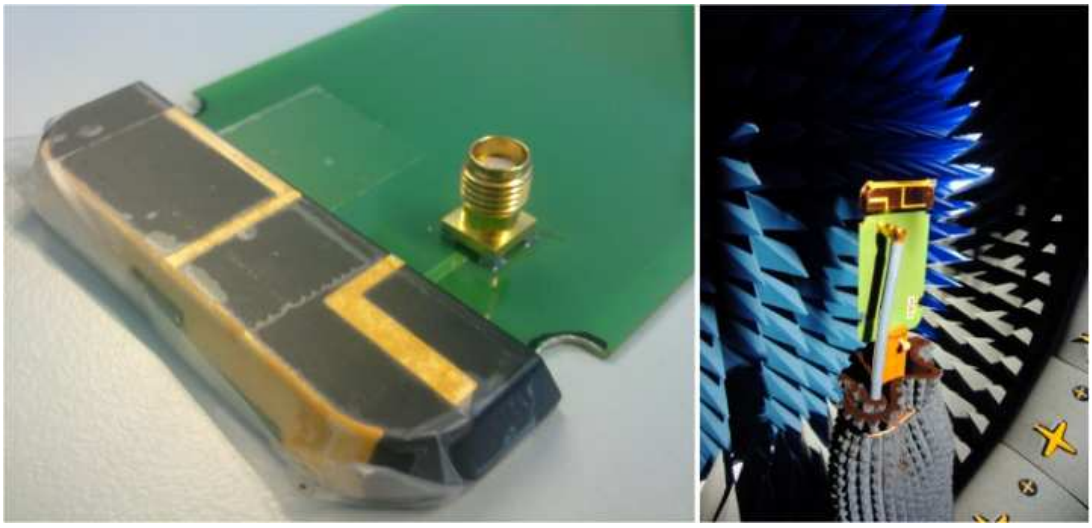

Figure 121 Manufactured LTE Antenna Prototype and Efficiency Measurement Environment

The s-parameter measurements for this antenna were performed also with the longer ground plane (90mm) configuration, whose results are presented in Figure 122. In Figure 123, an efficiency between approximately -7dB and -3dB is seen in the LB with short GND plane. However the efficiency drops further to -10dB for a narrow bandwidth, around 700MHz. When the ground plane length is increased to 90mm, an improvement is observed as expected, efficiency being between -5dB and -2.4dB. The measured efficiency is below -5dB between 700-800MHz for short ground plane and between 700-730MHz for long ground plane. The HB efficiency is not affected by the change in the length of the ground plane and is between -5dB and -1.8dB for both cases.

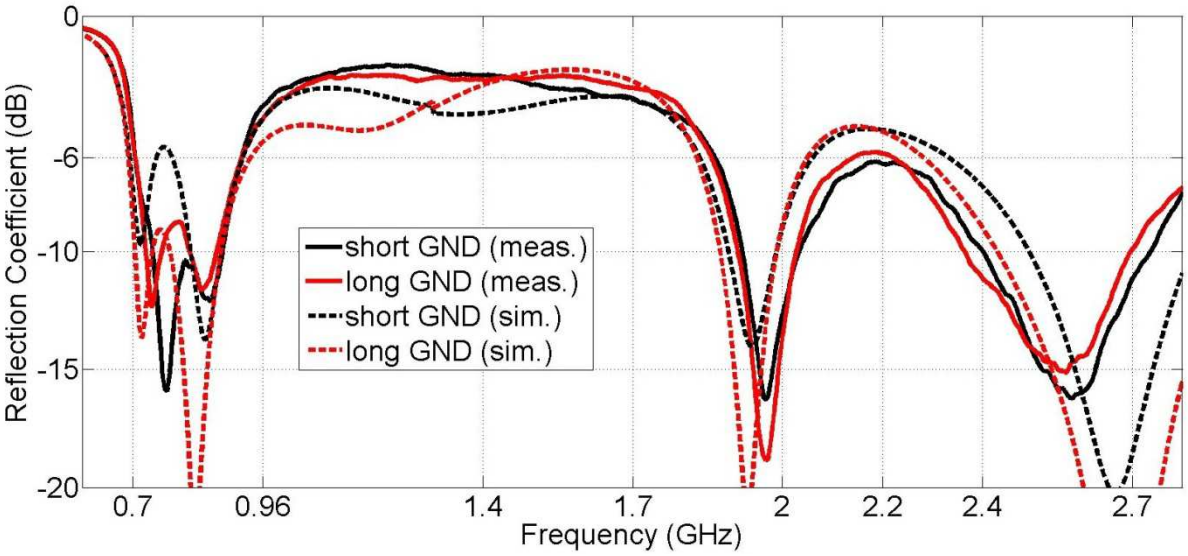

Figure 122 Measured/Simulated Reflection Coefficient of LTE Antenna with Short/Long Ground Plane

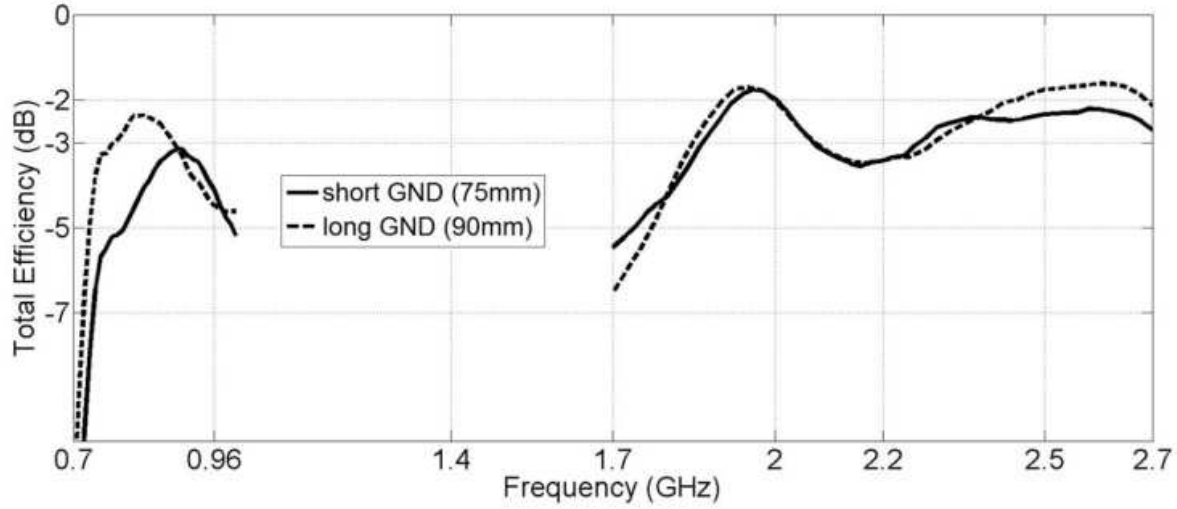

Figure 123 Measured Total Efficiency of LTE Antenna with Short/Long Ground Plane

To investigate the reason of the very low radiation efficiency at 700MHz, surface currents were plotted at 700MHz, 750MHz and 900MHz (Figure 124). The parasitic element is strongly excited at 700MHz and the current circulates following the path of: antenna feed-driven strip-parasitic strip-ground plane-back to the antenna feed. For this reason, the currents on the ground plane (especially on long edges of the PCB) are weak, when compared to the more uniform excitation (in the longitudinal direction) of the ground plane at 750MHz and 900MHz. Since most of the radiation comes from the first chassis wavemode at the lower frequencies, the low radiation efficiency can be explained by the currents circulating between the driven strip and parasitic strip and not forming the first wavemode.

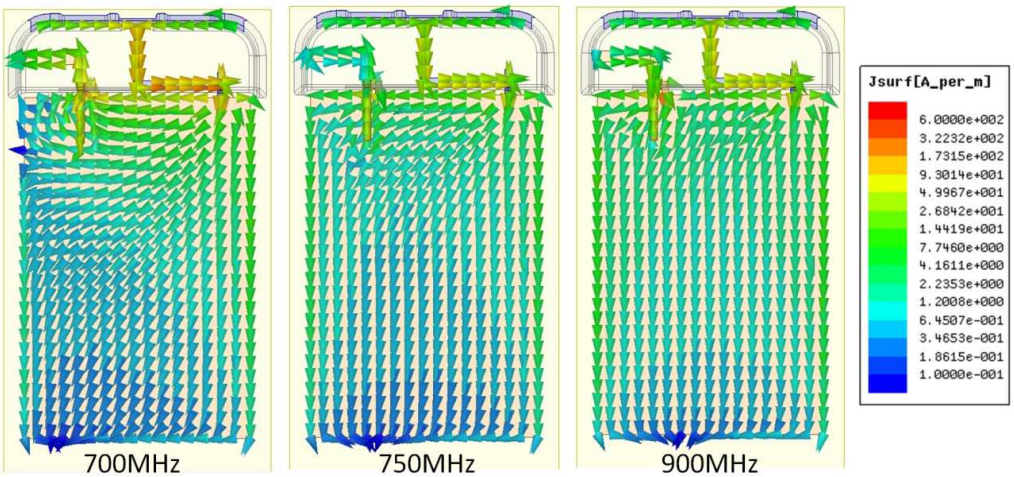

Figure 124 Surface Current Distribution of the LTE Antenna in the LB

3. INVESTIGATION OF THE USER EFFECT

The effect of the user's hand and head on the efficiency and matching of the antenna was investigated through measurements using the two antenna prototypes which were designed for free space. It is worthwhile to note that the antennas were not designed on purpose to be resistant to detuning and the main aim of the study was to form a basis for future studies.

As widely known, the user's hand and head affects the antenna performance in two ways. The first effect is the change in the reflection coefficient due to dielectric loading. The second effect is the reduced total efficiency, due to the lossy characteristics of different body tissues in the near field of the radiating elements and this last effect is difficult to avoid as a lot of power will be lost in those lossy tissues.

To investigate both these effects, s-parameter and total efficiency measurements were performed, using CTIA hand and head phantoms (Figure 125 on left taken from [7]). In order to perform a more realistic measurement, a dielectric casing made of ABS P430 material was used (white structure in Figure 125 on right). The relative permittivity of the ABS material is 2.4 and the loss tangent is 0.0053.

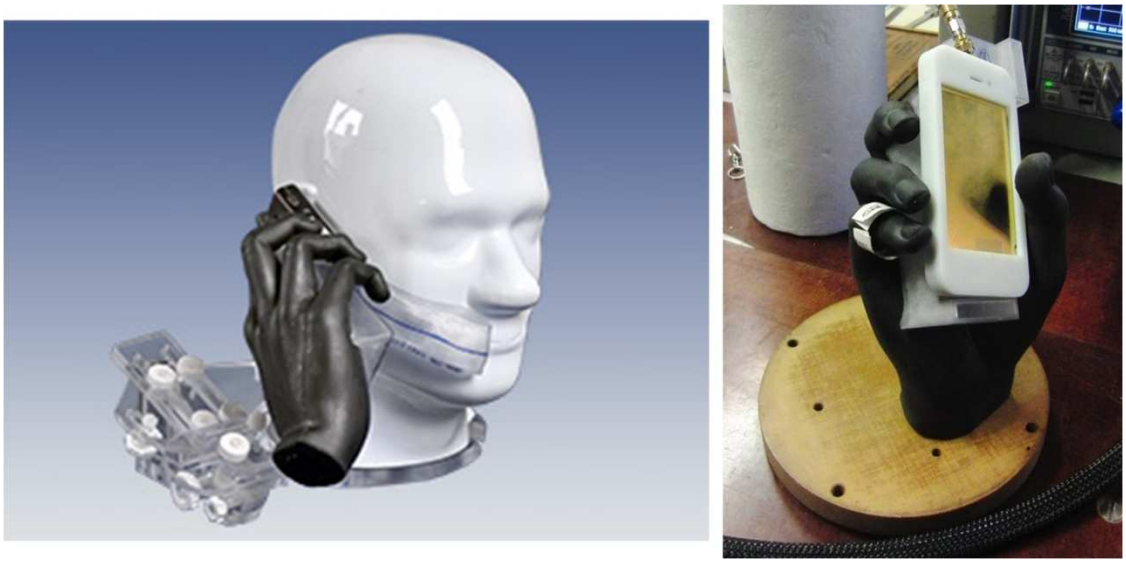

Figure 125 CTIA Hand and Head Phantoms and Dielectric Casing

Figure 126 Efficiency Measurement Setup Taking into Account the User's Head and Hand

The s-parameter measurements of the UMTS antenna in four different cases are presented in Figure 127. These four cases are defined as:

- Free Space
- Casing: The antenna placed inside the dielectric casing
- Hand: The antenna placed in the CTIA hand with the casing
- Head + Hand : The antenna with casing, placed between the CTIA hand and head phantom

It is necessary to note that the antenna was placed on the bottom of the PCB (facing the palm) in the measurements with hand, to stand for the real use case in today's mobile phones.

It is a well known fact that introducing the user's hand to an antenna will shift its operating frequency down, due to dielectric loading. This behaviour can be seen Figure 127 for the UMTS antenna. With the introduction of the hand and head into the s-parameter measurements, a shift in the frequency of minimum $|s_{11}|$ can be seen towards the lower frequencies. The detuning in the matching of the antenna is more dominant in the LB. The measured total efficiency for the four cases (free space, with casing, with casing and hand, with casing and hand and head) can be seen in Figure 128. It is observed that introducing the casing into the antenna had a positive effect on the efficiency through dielectric loading. The resonance created by the parasitic strip around 700MHz (which has lower efficiency than other frequencies since the currents are concentrated mainly on the parasitic strip so that chassis wavemode is excited weakly) is shifted down, which increases the radiation efficiency (and thus total efficiency) in the band of interest. When the hand phantom is taken into account, there is a decrease in the total efficiency of approximately 5dB in the LB, decreasing the efficiency to the levels below -8dB, due to both detuning effects and loss mechanisms. The same behaviour can also be seen in the HB, where the efficiency drops to the levels around -5dB due to losses in the hand. When the head is also taken into account with the hand, the antenna detuning gets more severe (in the LB) and further decrease is observed in the measured efficiency. The efficiency level is between -17dB and -14dB in the LB and between -13dB and -10dB in the HB.

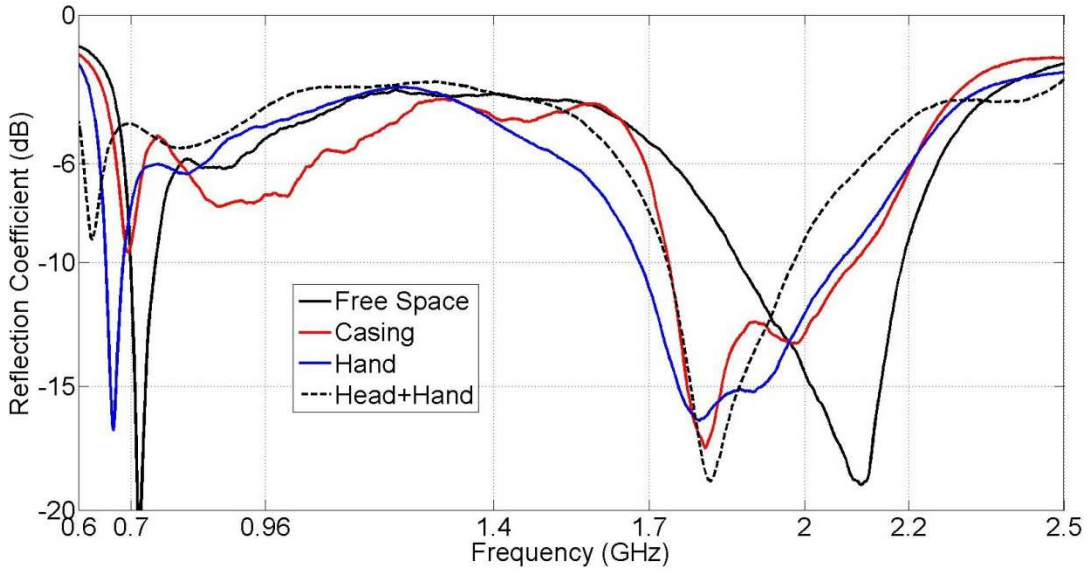

Figure 127 S-Parameter Measurements of the UMTS Antenna with User Effect

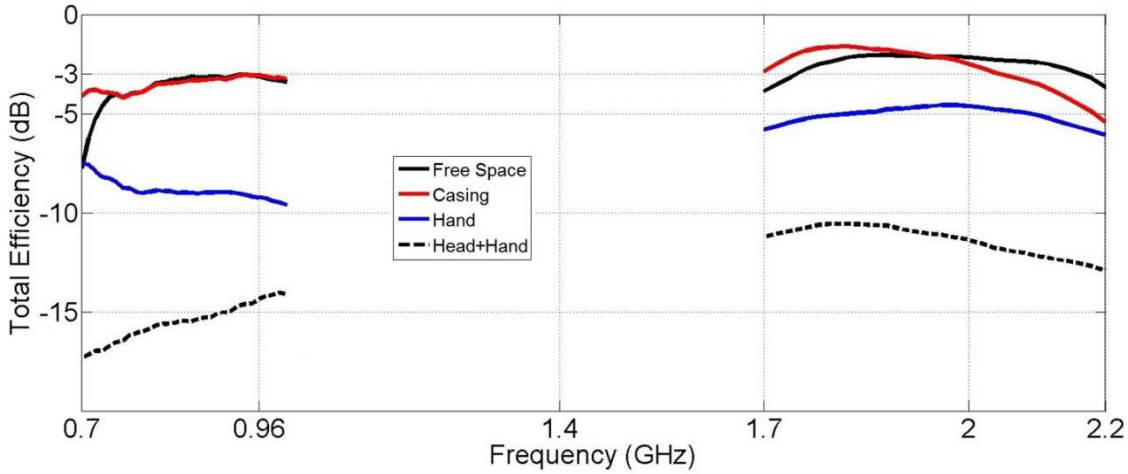

Figure 128 Total Efficiency Measurement of the UMTS Antenna with User Effect

The same comments are applicable also to the measurement results of the LTE antenna. The s-parameter and total efficiency measurement results for this antenna can be seen in Figure 129 and Figure 130 respectively. About the detuning, the antenna stays more or less tuned in the LB in "casing" and "hand" configuration except some narrow overshoots. The mismatch is severe when the head is introduced. The effect is less in the HB and the antenna is matched despite the user effect, although some overshoot is seen with the casing. For the efficiency, when the hand is taken into account, it drops to the levels between -10dB and -7.5dB in the LB and between -10dB and -3.5dB in the HB. When the head is also introduced, efficiency levels between -19dB and -17dB in the LB and between -13dB and -9dB in the HB are observed.

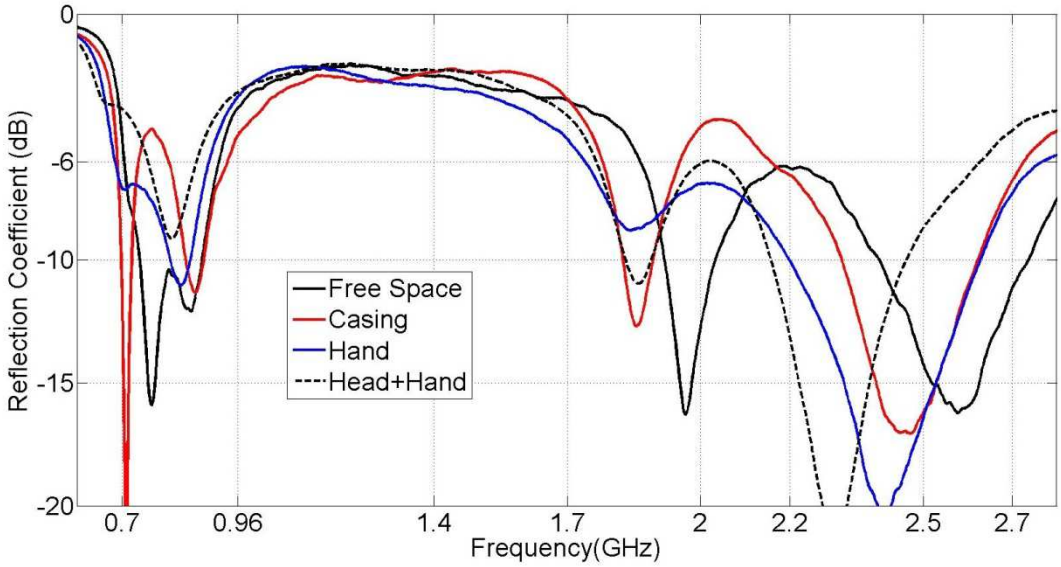

Figure 129 S-Parameter Measurement of the LTE Antenna with User Effect

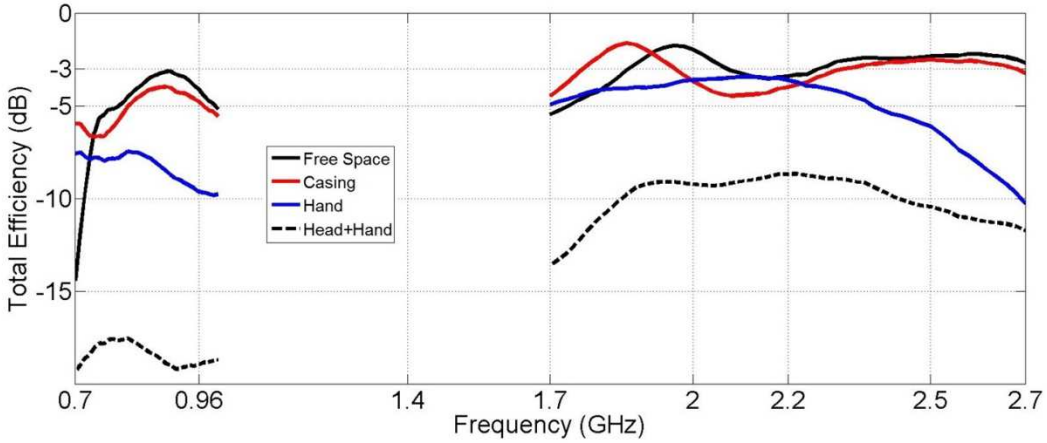

Figure 130 Total Efficiency Measurement of the LTE Antenna with User Effect

4. CONCLUSION

In this chapter, two different antenna designs were presented targeting the 4G communication frequency bands. Both antennas proposed are single feed antennas realized on plastic material using Laser Direct Structuring (LDS) technology. The major design challenge is the relatively short ground plane of 75mm, to rely only on the 3.5" touchscreen of a smart phone as the ground plane.

The first antenna consists of a driven strip, which capacitively excites a parasitic strip connected to the ground plane. With this topology, a high bandwidth potential can be achieved in the LB (700-960MHz) where the antenna is electrically small. With a four element MN introduced to the antenna feed, the low LTE (700-960MHz) and UMTS (1.7-2.2GHz) frequency range can be covered with a single feed design with a reflection coefficient below -6dB. The simulation results have been verified through measurements and total efficiency results for the antenna have been presented.

A similar topology was used for a second design where the driven strip capacitively excites a parasitic element but the orientation of the driven strip was changed. In this way, the additional coverage of the WLAN and high-LTE bands was also made possible.

Finally, an investigation about the effect of the user's hand and head was done using the two antennas. Measurements were performed with a dielectric casing, CTIA hand and head phantom to see the effect on both the s-parameters and total efficiency. Table 5 summarizes the measured total efficiency values in the presence of user hand and head. for the two prototypes. There is also a number of studies, investigating the effect of the user on antennas for mobile terminals like in [8-11]. For example in [4], a CE antenna with a shielding structure was proposed for GSM/UMTS bands. The shielding structure was used to reduce the interaction between the user and the antenna. In the "antenna on top" configuration, the measured efficiency was -10dB at 900MHz and -7.5dB in HB with hand and head. It is worthwhile to state that antenna on top configuration (when the antenna faces only the index finger) is more advantageous than "antenna on bottom" configuration used in this study since the detuning is less effective and absorption losses are less. In [9], a CE with multiband MN was used to cover low LTE/GSM/UMTS bands, with special attention to mitigate detuning in the design phase. The main target was to minimize the detuning especially in the LB when the antenna is placed on top of the PCB. The measurements were performed with the hand only in both on top and on bottom placements. An efficiency drop of approximately 7dB was seen in efficiency measurements with hand when compared to free space. In [10], a dual feed antenna structure was proposed on Rogers 4003 substrate. The LB coverage (700-960MHz) was achieved thanks to a CE and the HB coverage (1.7-2.7GHz) was done using a portion of the CE by offset feeding. This antenna was also designed on purpose to be resistant to detuning when the antenna is on top of the PCB. The measured efficiency drop (compared to free space) with the hand (when antenna is on bottom) was -6.5dB in LB and -7dB in HB. Finally in [11], the user effect was investigated on a narrowband PIFA antenna covering GSM900/DCS1800 frequency bands. An efficiency drop of 11dB at 900MHz and 8.5dB at 1800MHz was measured with hand and head when the antenna is placed on bottom of the PCB.

When the data in Table 5 are compared with corresponding results from these studies, it can be concluded that the efficiency levels obtained by the two antennas presented here are competitive in terms of total efficiency. It should also be noted that no specific effort was spent considering the user effect in the design phase, and a ground plane of only 75mm was used in this study. It was previously shown that an additional increase of 0.5dB can be obtained making the ground plane length 90mm.

Table 5 Summary of the Measured Efficiency Taking into Account the User Effects

		UMTS Antenna	LTE Antenna
With Hand	Eff. in LB	Between -7dB & -9dB	Between -8dB & -10dB
	Eff. Drop from FS (LB)	4dB	3dB
	Eff. in HB	Between -4dB & -6dB	Between -10dB & -3dB
	Eff. Drop from FS (HB)	3dB	2dB
With Hand+Head	Eff. in LB	Between -17dB & -14dB	Between -19dB & -17dB
	Eff. Drop from FS (LB)	11dB	14dB
	Eff. in HB	Between -10dB & -13dB	Between -13dB & -9dB
	Eff. Drop from FS (HB)	9dB	7dB

5. PERSPECTIVES

It was mentioned previously that the results obtained from user effect study will be used for future studies on this subject. As also mentioned in Conclusion part, there has been a number of studies to design mobile terminal antennas resistant to detuning (mismatching) when the antenna is located on the top of the PCB. This configuration is less problematic when compared to "antenna on bottom" configuration, since the antenna only faces the index finger (instead of the whole palm) which results in less severe detuning compared to bottom location. It is also worthwhile to note that the antenna is generally located at the bottom of today's smart phones. For these reasons, it would be better to focus the future studies on an antenna resistant to detuning due to user's hand and head, when the antenna is placed on the bottom of the PCB.

In addition to the antennas presented in this chapter, some more measurements were also performed on the hollow CE antennas (presented in chapter III) with the CTIA hand. The main target was first to obtain the general behaviour of the impedance change of an optimally-overcoupled antenna(placed on the bottom of PCB) in presence of the user's hand. The measured input impedances of the two different hollow CE antennas are given in Figure 131 for free space (FS) and with hand configurations.

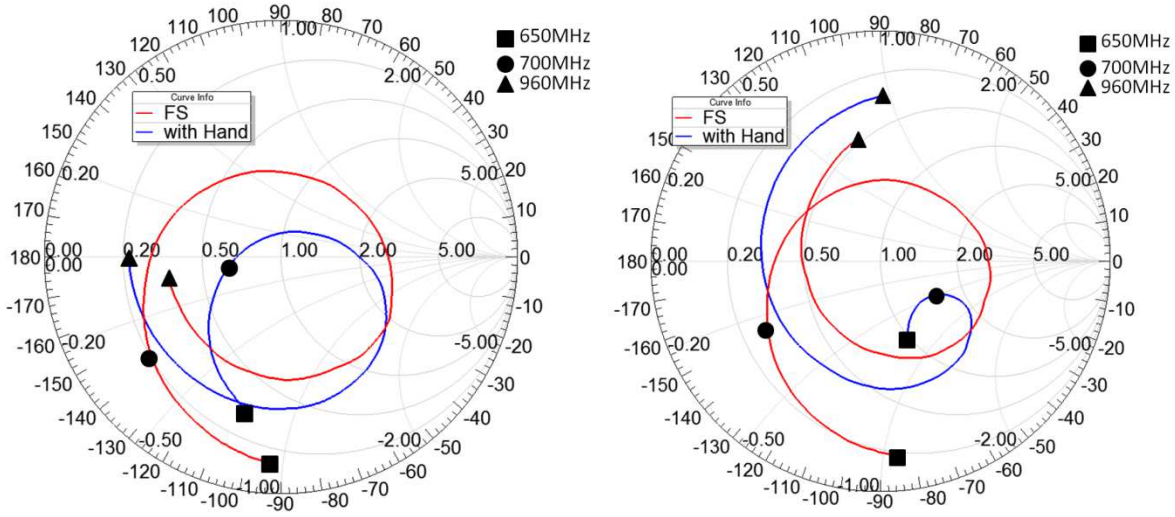

Figure 131 Measured Input Impedance of Hollow CE Antennas in FS and with Hand

Evaluating these results about the input impedance in a geometrical way in smith chart, it can be told that the impedance loop in FS condition is subject to shrinking when the user's hand is introduced. Due to the dielectric loading, the lower frequency point of the loop is pulled through the middle, where the higher frequency points are pushed out of the loop (Figure 132).

This can be verified by turning back to the results in Figure 131. For the antenna on left, the 700MHz point (shown with a circle) is initially quite outside of the starting region of the loop (in FS). When the hand is introduced, this point goes inside the loop and as a consequence the starting frequency of the loop falls to frequencies like 650MHz. The 960MHz point (shown with a triangle) which is close to the end point of the loop in FS, is pushed out of it when the hand in introduced. A similar behaviour can also be seen for the results in the smith chart on right.

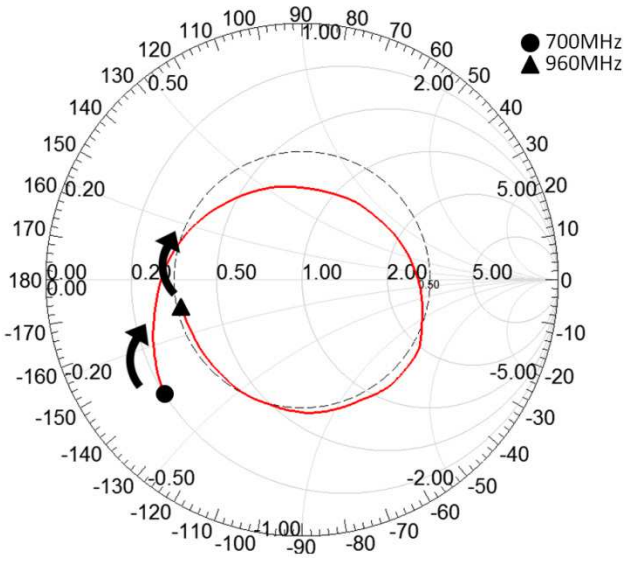

Figure 132 Geometrical Analysis of the Impedance Change with User's Hand

Using this background, initial studies were done to design an antenna resistant to detuning due to hand. The model of this initial design can be observed in Figure 133. The antenna is a dual-feed structure where the first feed is responsible for covering the LB (700-960MHz) and second feed for HB coverage (1.7-2.7GHz). The LB antenna consists of a short driven strip which capacitively excites a parasitic strip connected to ground plane through an inductor. A series inductor is used at the LB feed for impedance matching. The HB antenna is a separately fed monopole, with a two element MN at the feed. The antenna is totally printed on an FR4 substrate having dimensions of 130X60X0.8mm^3. A ground clearance region of 10mm is reserved for the antenna, making the ground plane 120mm long. The simulations of this antenna was performed in a commercial full wave electromagnetic solver, EMPIRE XCcel [12], in free space, with hand and with hand+head conditions(Figure 134).

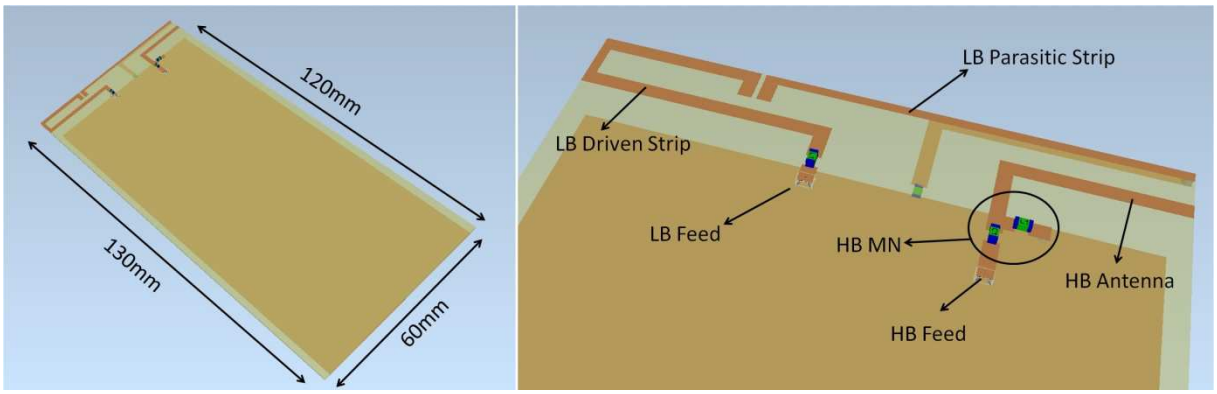

Figure 133 Simulation Model of the Antenna Resistant to Detuning

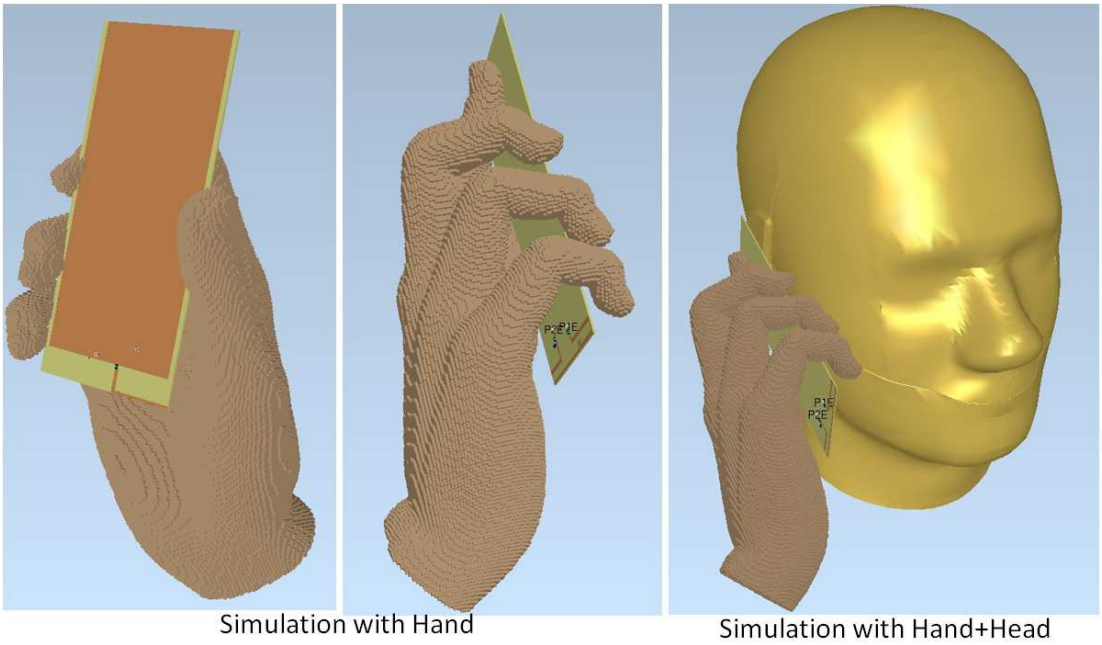

Figure 134 Simulation Models of the Antenna Considering the User

The simulated S-Parameters of the antenna in FS and "with Hand" conditions are given in Figure 135 and Figure 136. The antenna can cover the target bands with a reflection coefficient below -6dB in both FS and "with hand" conditions. In the LB, the matching bandwidth is reduced from 390MHz to 350MHz when the hand is introduced but this bandwidth is still enough to cover the LB.

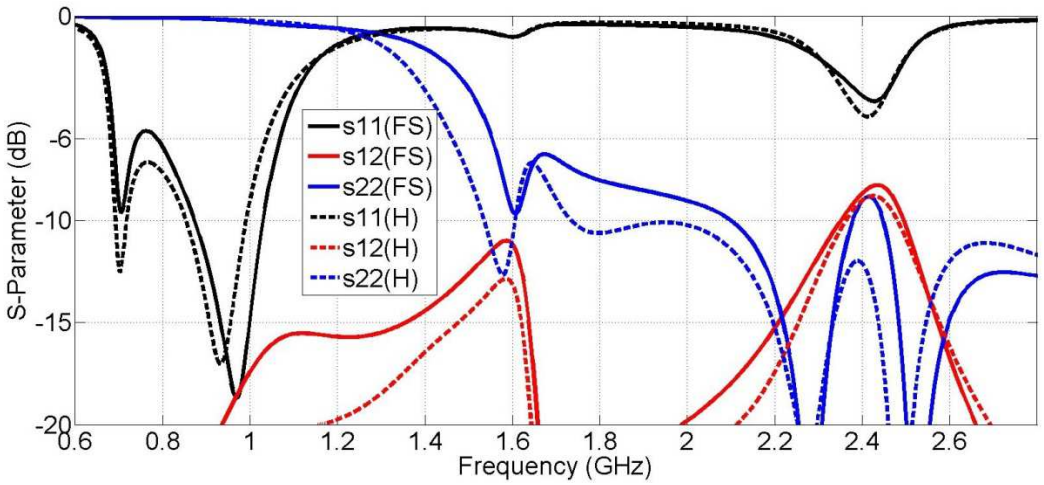

Figure 135 Simulated S-Parameters of the Antenna in FS and with Hand

The effect of the hand on the input impedance in LB occurs as foreseen. The loop in the smith chart shrinks and the high edge frequencies of the impedance are pushed out of the loop. Special attention was paid for situating the 960MHz point, such that it will not get out of the VSWR=3 circle so it will still be matched even with the hand. When the head is also taken into account in simulations, the antenna continues being matched in 700-960MHz with a reflection coefficient below -6dB(Figure 137).

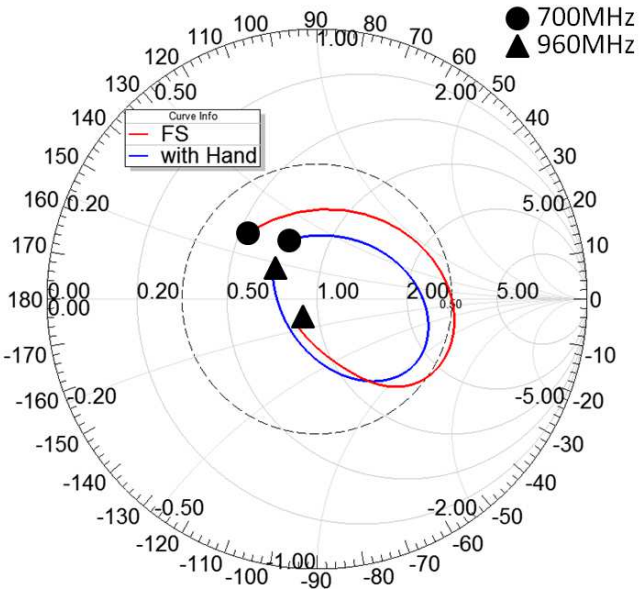

Figure 136 Simulated Input Impedance of the Antenna in FS and with Hand (Smith Chart)

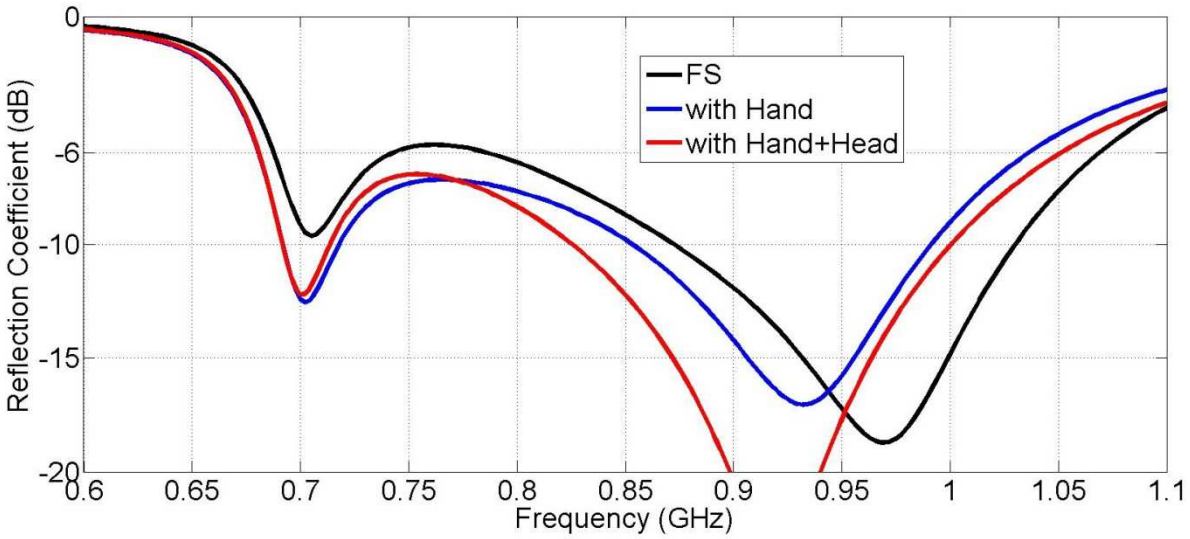

Figure 137 Simulated S-Parameters of the Antenna in LB

The proposed antenna has been manufactured. Figure 138 and Figure 139 presents a comparison of the simulated and measured S-Parameters of the antenna in FS. A good compliance can be seen between the simulations and measurements, except an overshoot in the LB, between 750-830MHz rising up to -4.5dB. However it can be told from the LB input impedance in Figure 139 that this overshoot is not something about the bandwidth potential, it can easily be overcome by tuning the component values in the LB MN.

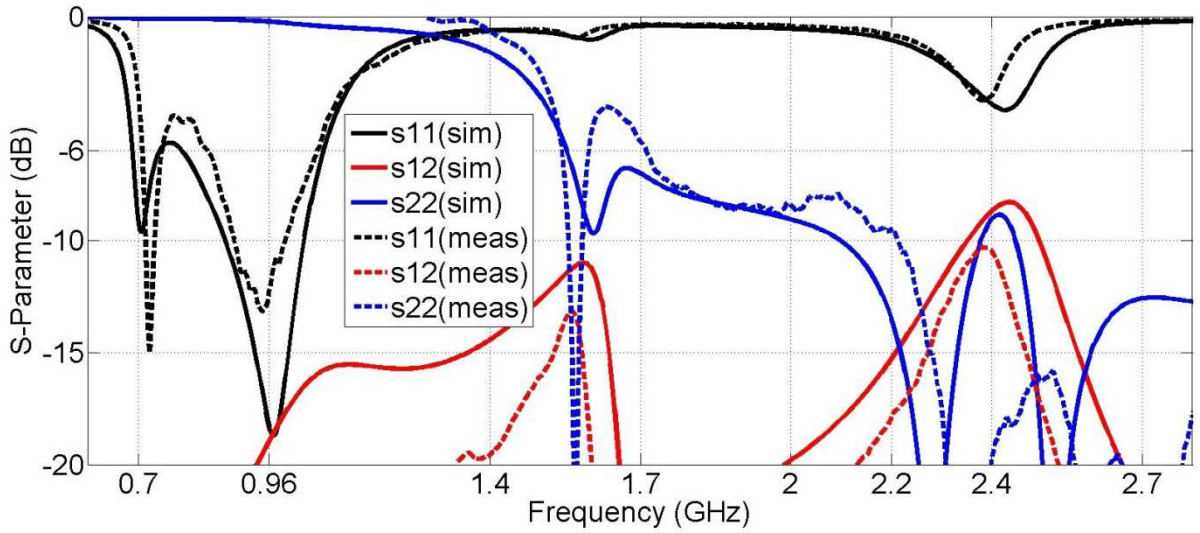

Figure 138 Simulated and Measured S-Parameters of the Antenna in FS

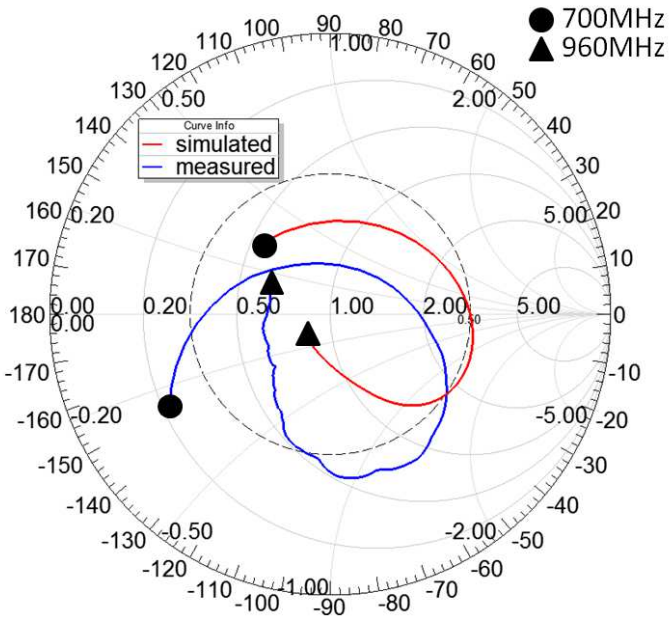

Figure 139 Simulated and Measured Input Impedance (LB Only) of the Antenna in FS (Smith Chart)

S-Parameter measurements were also performed for this manufactured prototype, using a real human hand in talking grip position when the antenna is on bottom of the PCB. The measured S-Parameters can be seen Figure 140 in dB format and also in Figure 141 for the Lb in smith chart. The antenna keeps its reflection coefficient below -6dB between 700-960MHz with LB antenna and between 1.7-2.7GHz with HB antenna, in the presence of a real user hand.

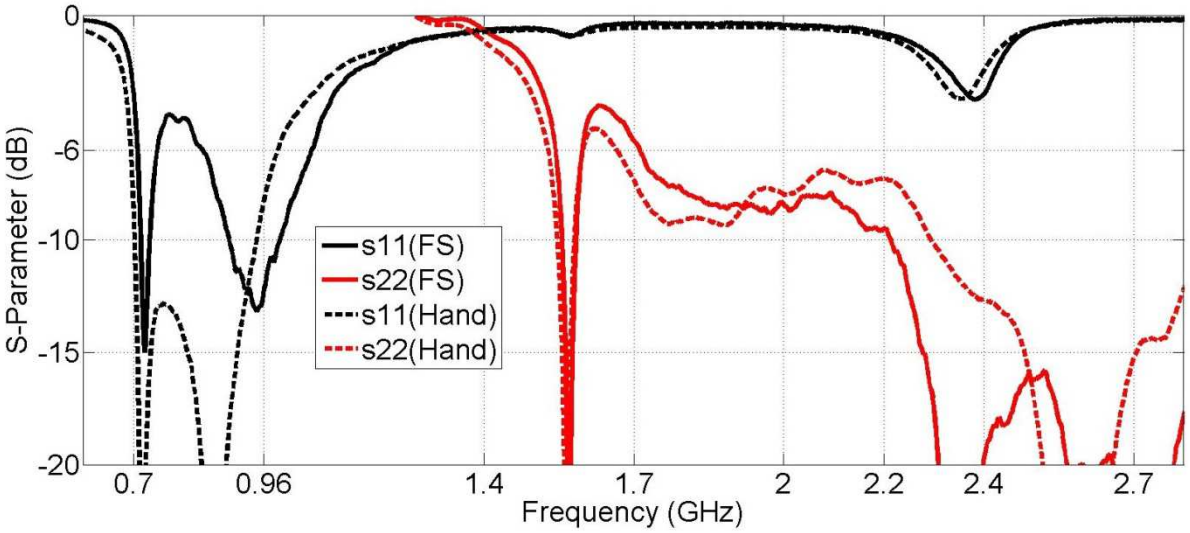

Figure 140 Measured S-Parameters of the Antenna in FS and with Hand

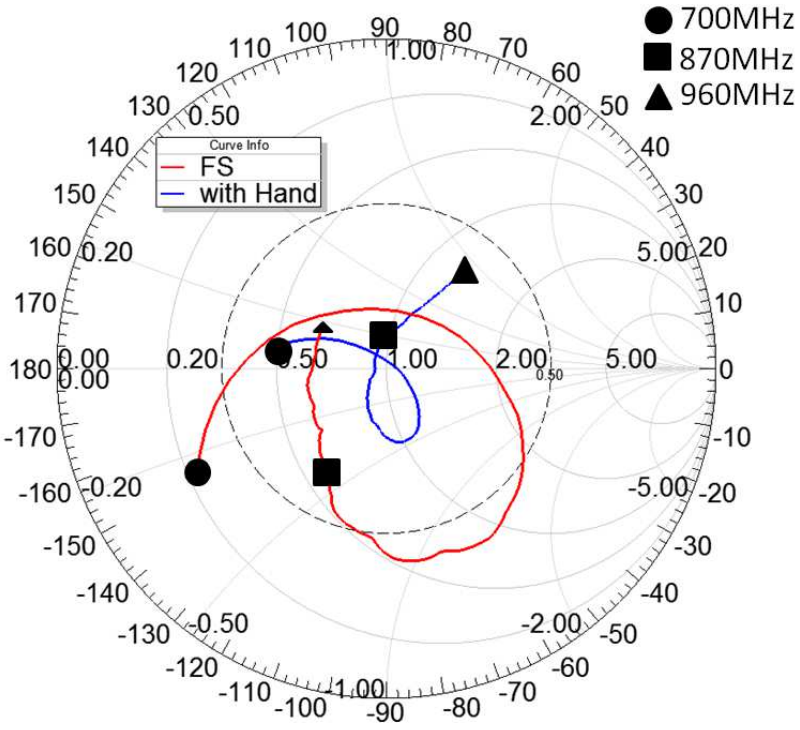

Figure 141 Measured Input Impedance of the Antenna (in LB) in FS and with Hand (Smith Chart)

For future studies, it is targeted to take this antenna as a basis and improve the design for better performance in the LB in FS condition. The S-Parameter and efficiency measurements are also next steps for validation of the concept.

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CHAPTER V
EYEWEAR ANTENNAS

This section presents some eyewear antenna concept that can cover 700-960MHz and 1.7-2.7GHz, to be used on wirelessly connected eyewear devices. Since the antennas for eyewear devices operating in 4G cellular communication standard is a new subject, firstly a feasibility study is done for the bandwidth potential of the antenna in different placement alternatives. In all the study, coupling element (CE) antennas totally printed on FR4 substrates are used. To obtain a good compliance with the major use cases, the human head is always taken into account in simulations. Afterwards, three antenna designs are proposed, with a proper dielectric casing and frame. The SAR simulations are also performed and the effect of the user hand is also investigated.

1. USE CASE AND GENERAL INFORMATION ON EYEWEAR DEVICES

Wearable devices incorporating wireless connection with the peripherals has gained a significant popularity in the last years. Some examples of such devices can be considered as the wireless keyboard and mouse, cameras, USB keys, etc. In the recent years, wearable accessories like watches, rings and glasses have also gained importance as communicating devices. One of the most popular of these devices is the eyewears (or glasses), that currently has Wi-Fi 802.11b/g or Bluetooth connection with a peripheral mobile phone or a hotspot. There are currently many ongoing projects by different companies, who target the release of these "smart eyewear" devices in 2014 like [1-5]. Some examples of these devices are presented in Figure 142. Another example of such a device [1] can be seen in Figure 143. This device has a mechanical structure that can be integrated with any glass frame especially if it is optical or solar. The battery is placed behind the user's ear, together with the Wi-Fi/Bluetooth antenna. On the side of the head, a board is found which incorporates the electronics like the touchscreen and video/audio processors. The connection of the transceiver and the antenna is done using an RF cable passing through the plastic frame of the glasses. A camera is placed in the region corresponding to the front of the head. The display issue is solved by using a prism lens in front of the eye, in a way with no disturbances to the actual sight of the eye. A photo showing the positioning of the electronics in the eyewear device is given in Figure 144.

Figure 142 Different Eyewear Devices under Development

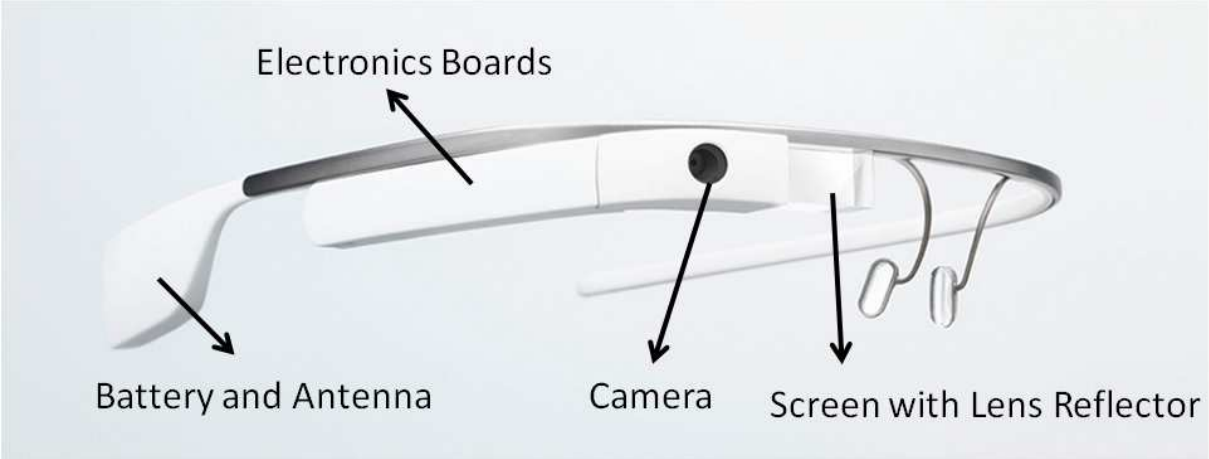

Figure 143 A Generic Eyewear Device Example

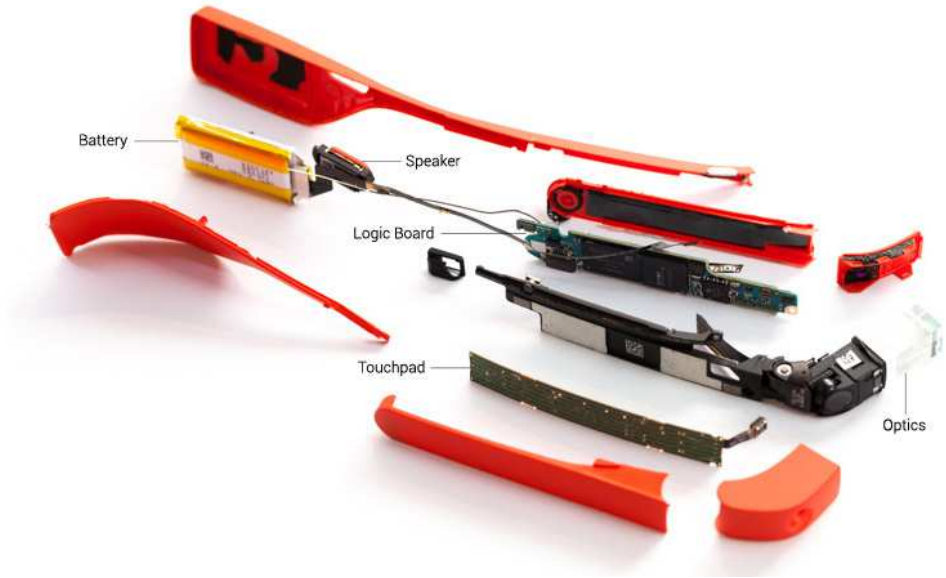

Figure 144 Teardown of the Google Glass

For now, the device can accept calls or use navigation systems by connecting to a mobile phone using Bluetooth. It also allows internet navigation if connected to a hot-spot. The antenna for this connection is placed behind the ear similar to headsets for mobile phones. This placement is practical for now, since the antennas of the same type with similar functionality were investigated extensively in the past in Bluetooth headsets[6-7]. The antenna design problem is relatively easy to solve since the space for the antenna is comparable to the wavelength at this frequency. Also since the power levels for Bluetooth are very small (generally 2.5mW), the SAR is out of concern and no regulation exists[8-9] However it is widely foreseen that these devices may replace mobile phones in the upcoming years. So, there is clearly a need for 4G cellular antennas for these eyewear devices, with satisfactory performance and compliance with the SAR requirements in the head of the user.

2. FEASIBILITY STUDY FOR EYEWEAR ANTENNAS

As an application example of coupling element (CE) antennas on eyewear devices, a feasibility study was first performed (simulations) taking into account a homogeneous user head. In the

simulations, CEs which are totally printed on the 0.8mm thick FR4 substrate were used in three different antenna locations (spotted to be available) and the bandwidth potentials and SAR values in these three locations were compared, using the commercial software EMPIRE XCcel [10]. Figure 145 shows the general model used in the simulations. The PCB that will support the antenna is modeled as FR4 substrate of 0.8mm thickness and with the dimensions given in Figure 146 (chosen as realistic dimensions versus a human head). A 90° bend is added to the PCB to stand for the region that extends in front of the eye. The CE is printed on the inner side of the PCB and the ground plane is on the outer side of the substrate.

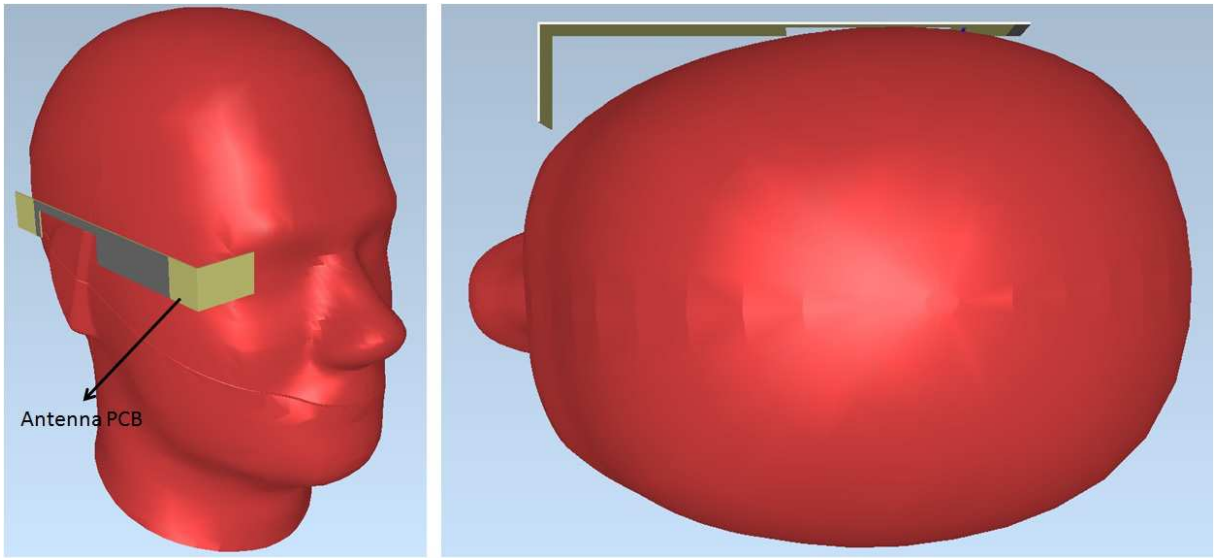

Figure 145 Simulation Setup for the Feasibility Study (3D view and top view)

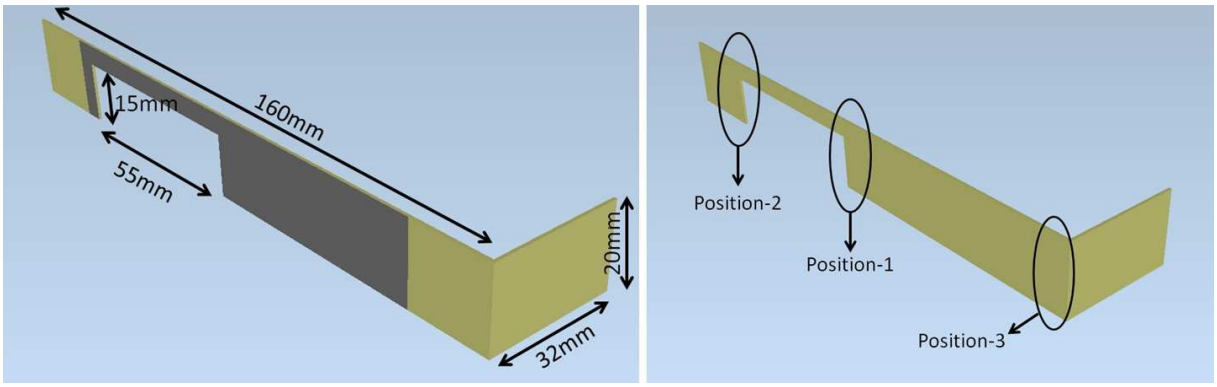

Figure 146 PCB Dimensions and Evaluated Antenna Positions

Three different antenna positions were evaluated in terms of bandwidth potential. These locations are: behind the ear, in front of the ear and close to the eye, as shown in Figure 146.

2.1. INVESTIGATION OF BANDWIDTH POTENTIAL IN POSITION-1

The simulated bandwidth potential of a basic printed CE (in the presence of the user head) in position 1 (in front of the ear) is presented in Figure 147. The bandwidth potential is computed using the Optenni Lab software using a two element MN at each frequency point as described in more detail in previous chapters, so it is an obtainable realistic value compared to theoretical values. To

cover the 4G frequency spectrum, a bandwidth potential of approximately 260MHz at 830MHz and a potential of 1GHz at 2.2GHz is necessary. It is worthwhile to note that these bandwidth potential values are the necessary values for single band coverage, with a two element MN. However since a dual-band coverage is necessary here, it was targeted to get "more than enough" bandwidth potential in the HB at the same time with enough potential in the LB. If Figure 147 is observed, it can be noted that the bandwidth potential may not be sufficient to achieve dual-band coverage (700-960MHz and 1.7-2.7GHz) with a simple MN, since the potentials in the LB and HB do not seem to be high enough. For this reason, this CE in position-1 was evaluated covering only the 1.7-2.7GHz with the introduction of a 2 element MN for SAR simulations. The simulated reflection coefficient with the MN can be seen in Figure 148.

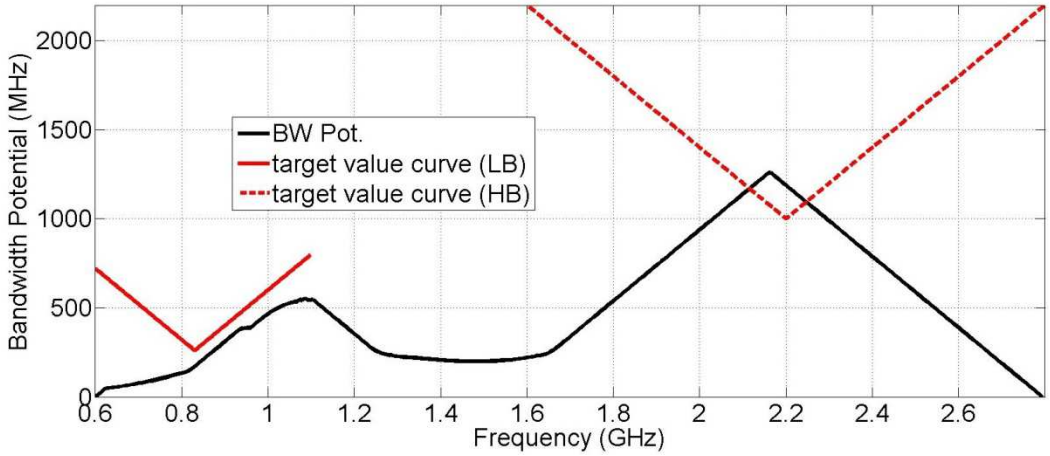

Figure 147 Bandwidth Potential for Position-1

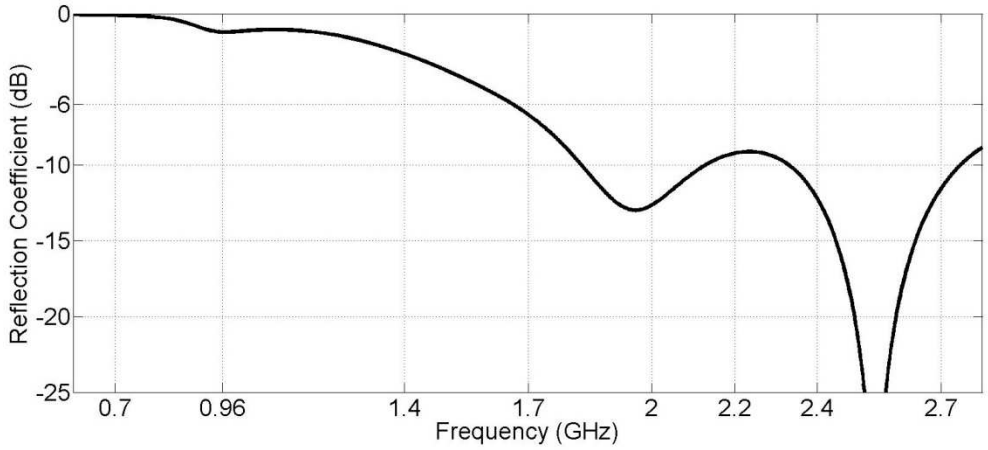

Figure 148 Reflection Coefficient for Position-1

The antenna has a radiation efficiency varying between -7.2dB and -4.4dB in 1700-2700MHz frequency range. It can be noticed that the efficiency of the eyewear antenna is considerably higher than the efficiency of mobile terminal antennas (in chapter IV) in hand+head condition, which was on average -10dB in HB. The main reason is the unnecessity of the user hand in the eyewear usage scenario, which enables getting rid of the losses due to detuning and absorption in the hand. The 3D gain pattern of the antenna at 2200MHz is presented in Figure 149. The radiation is concentrated through the direction away from the head. The gain is lower in the direction of the head due to absorption as expected. When the PCB is held in horizontal direction, the horizontal polarization is

dominant than vertical at 2200MHz with a small difference (0.7dB) between the gains of cross polarizations.

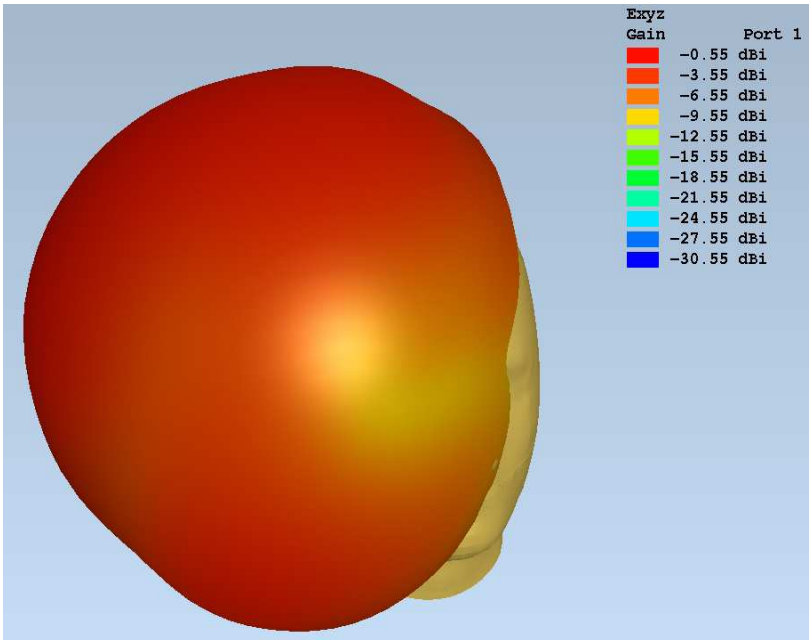

Figure 149 3D Gain Pattern for Position 1 (2200MHz)

2.2. INVESTIGATION OF BANDWIDTH POTENTIAL IN POSITION-2

The bandwidth potential for the CE printed on the substrate in position-2 (behind the ear) is presented in Figure 150. Placing the antenna in this position significantly has the advantage of higher bandwidth potential, sufficient for simultaneous coverage of LB and HB. The main reason of this higher BW potential especially in the LB is the stronger excitation of the ground plane in this position, when compared to previous one. A MN of two elements was inserted at the feed of the antenna and the reflection coefficient in Figure 151 was obtained for SAR simulations.

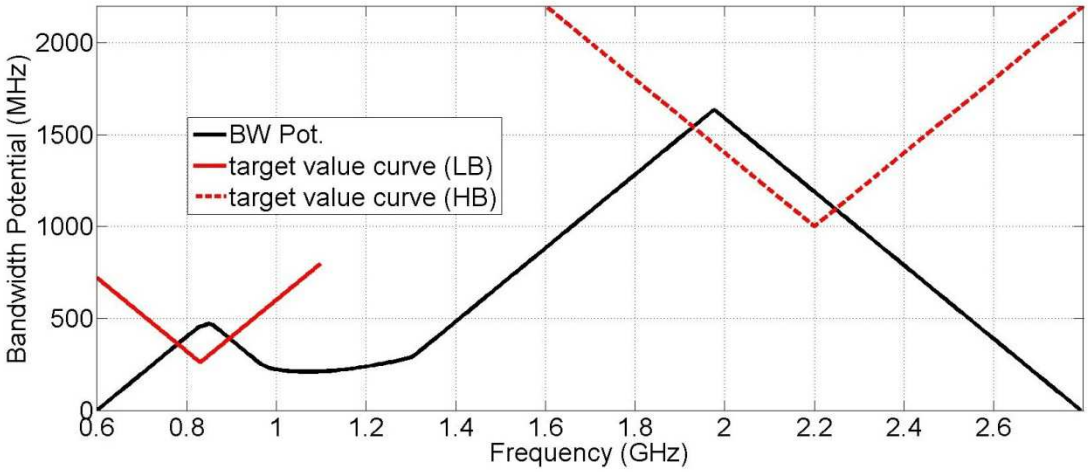

Figure 150 Bandwidth Potential for Position-2

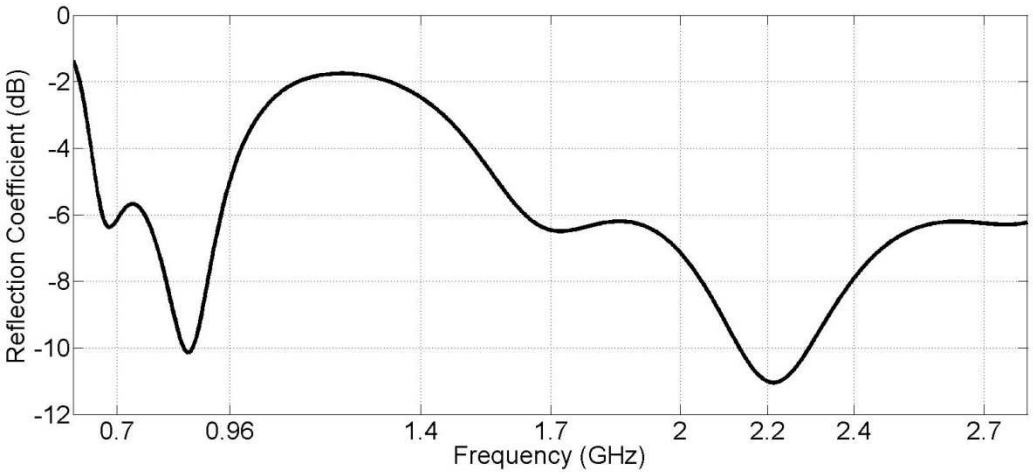

Figure 151 Reflection Coefficient for Position-2

The simulated radiation efficiency is between -11dB and -9.7dB in LB and approximately -7dB in the HB. As explained before, the radiation efficiencies are considerably higher than the general efficiencies of mobile terminal antennas (-14dB in LB and -10dB in HB for antennas in Chapter IV). The 3D patterns can be observed in Figure 152 for two different frequencies, showing the radiation away from the head. The cross polarization level is about 9dB in LB and 5dB in HB where the horizontal polarization is dominant.

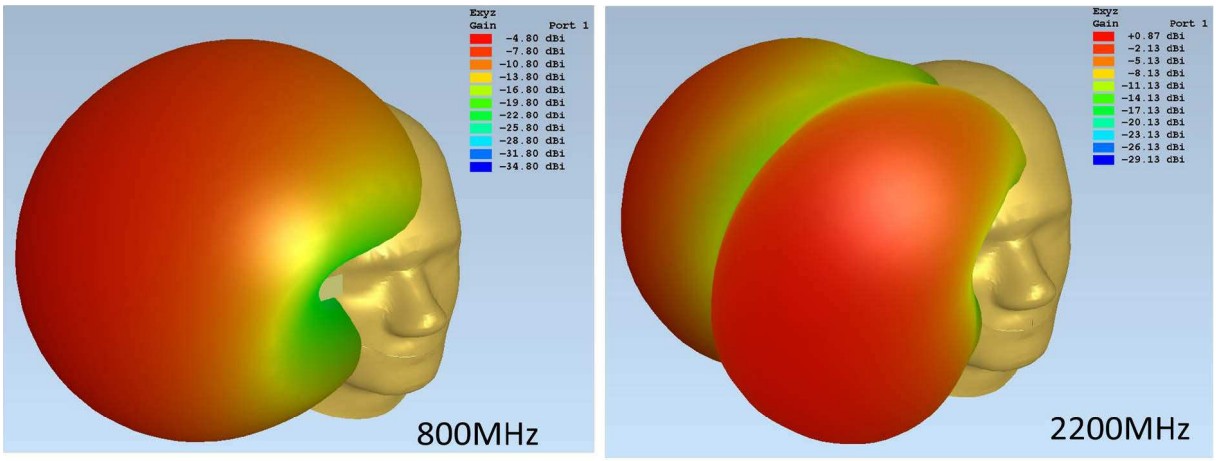

Figure 152 3D Gain Pattern for Position 2

2.3. INVESTIGATION OF BANDWIDTH POTENTIAL IN POSITION-3

The simulated bandwidth potential and the reflection coefficient of the antenna (with a three element MN) in position-3 (close to the eye) are given in Figure 153 and Figure 154 respectively.

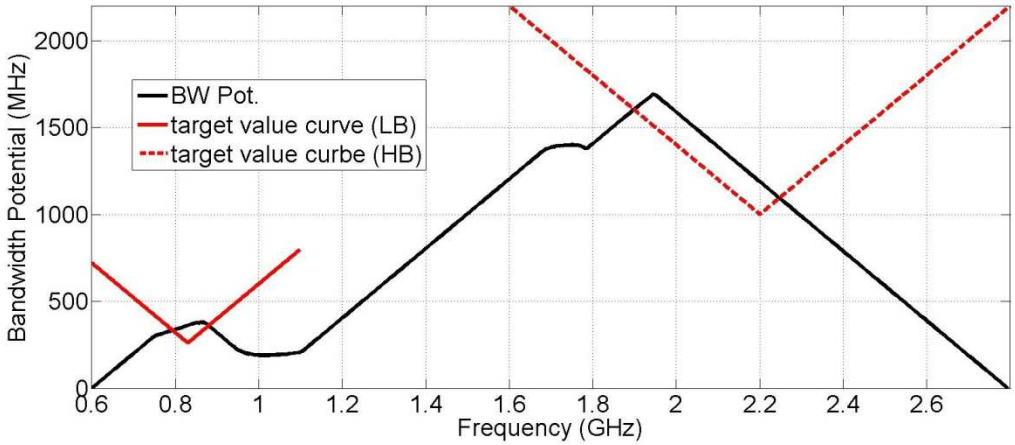

Figure 153 Bandwidth Potential for Position-3

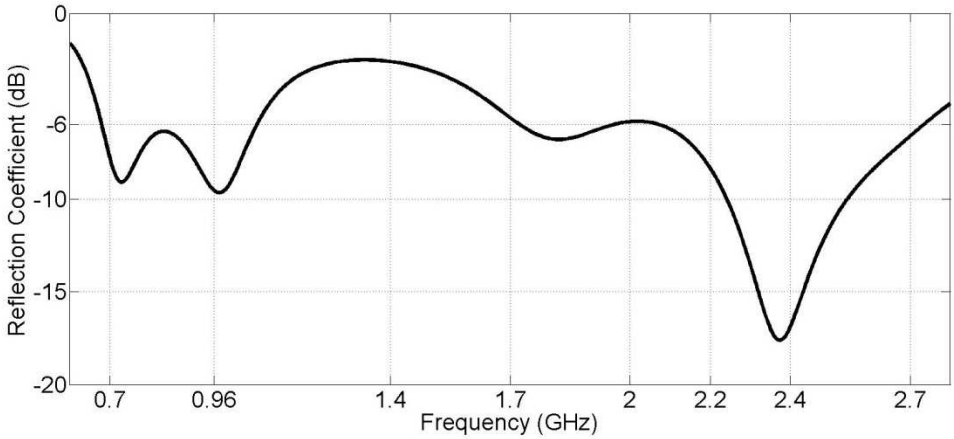

Figure 154 Reflection Coefficient for Position-3

The radiation efficiency is simulated to be between -9dB and -8dB in the LB and between -4.8dB and -3.8dB in the HB. The efficiency values seen in this position are quite higher than previous positions since the CE is located further away from the head when compared to others. The 3D patterns are presented in Figure 155. The horizontal polarization is generally dominant with a cross polarization of 6dB in LB and 4dB in HB.

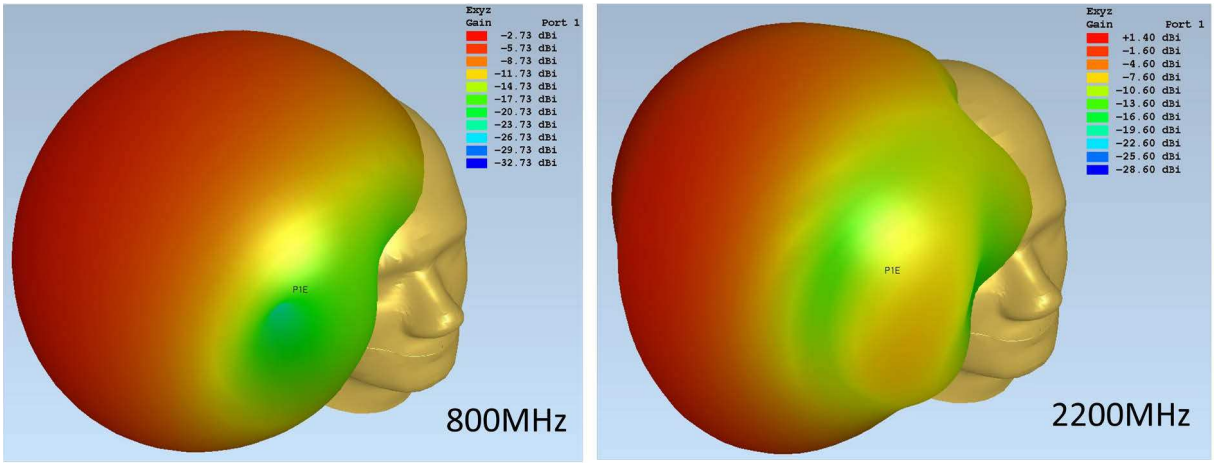

Figure 155 3D Gain Pattern for Position 3

2.4. COMPARISON OF SAR PERFORMANCES IN THREE LOCATIONS

Using the three CEs with associated MNs in the three different locations, a comparison about the SAR was performed. The simulation results are summarized in Table 6. The SAR results are presented both for the 1g SAR standard [11] (maximum limit 1.6W/kg, accepted in U.S) and 10g SAR standard [12] (maximum limit 2W/kg, accepted in Europe). The results are given also as normalized to 0.25W incident power (maximum of 2W in 1/8 second) and 0.25W antenna input after (at an incident power level to obtain 0.25W after mismatch). The SAR values that are below the allowed limit are shown in bold in the table. It can be seen that the SAR values are higher than the SAR values found from the measurements of a typical mobile phone, due to the proximity of the antenna feed to the head, the lack of a shielding ground plane under the antenna and also the lack of a dielectric casing material between the antenna and the head. It is also worthwhile to note that the radiation efficiency values are higher than a typical mobile phone, since for a mobile terminal, the antenna is covered by the hand in one direction and by the head in the other direction, which increases the losses due to absorption and mismatching. Another result to note is the lower SAR value of the location-3 in comparison to the other two locations, since the distance of the antenna element to the tissue is larger. As a conclusion, it can be told that the simulated 1g SAR values are above the standard limit, whereas 10g SAR values are generally acceptable or close to the limit.

Table 6 SAR Comparison for Eyewear Antenna Locations

			0.25W incident power		0.25W power after mismatch	
Position	Freq (MHz)	Radiation Eff (%)	1g SAR	10g SAR	1g SAR	10g SAR
1	1900	25	7.86	3.32	8.37	3.53
1	2200	36	4.31	2.07	4.88	2.35
1	2500	42	4.36	**1.54**	4.39	**1.55**
2	835	10	4.33	2.08	4.96	2.39
2	1900	20	2.97	**1.47**	3.89	**1.92**
2	2200	21	4.18	**1.99**	4.54	2.16
2	2500	21	3.68	**1.83**	4.69	2.33
3	835	13	4.17	2.03	5.50	2.68
3	1900	38	2.47	**1.18**	3.19	**1.53**
3	2200	33	2.22	**1.04**	2.60	**1.22**
3	2500	37	1.96	**0.82**	2.11	**0.89**
Limit			1.6	2	1.6	2

It is well known that the situation with a real head is different from the case with SAM phantom, since the real human head is composed of different tissues at different locations of the head. It was demonstrated in [13-15], that different parts of the body (for example the eyes) may resonate where the fields tend to increase and do not decay with distance. For this reason, a collaboration was established with Loughborough University to investigate the SAR with a visible human (VH) phantom in simulations (Figure 156). Comparisons using simulation results with different human head models were done in [16-17]. Here, the SAR simulation results with VH phantom give similar results with the simulation results using SAM phantom, except minor differences, due to small changes in the distance of the tissues to the head. It is observed that for all antenna positions and frequencies, the maximum SAR is observed on the side of the head. However, for antenna positions 1 and 3, secondary hot spots can be seen around the eye.

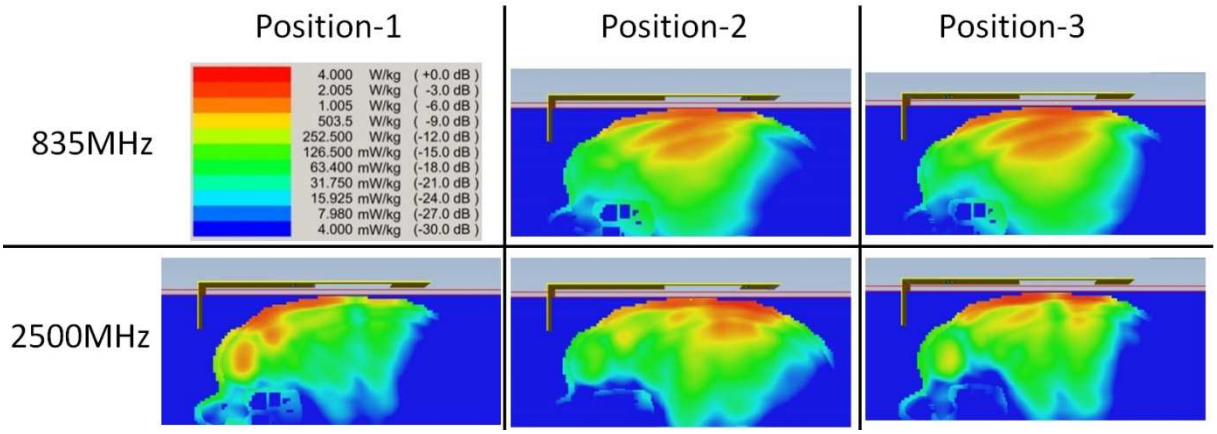

Figure 156 Simulated SAR with the Visible Human Phantom

3. EYEWEAR ANTENNA SOLUTIONS FOR 4G COVERAGE

Keeping in mind the results of the feasibility study, it was foreseen as possible to use printed CEs with corresponding matching networks, to obtain coverage of the 700-960MHz and 1.7-2.7GHz for 4G cellular communication standard. For this purpose three different CE antennas are proposed in this section. With the first antenna, the band between 700-2700MHz is continuously covered (including also the GPS operating frequency band), whereas the other two antennas can cover 700-960MHz and 1.7-2.7GHz simultaneously.

For a more realistic scenario, a dielectric casing with relative permittivity of 2.97 and a loss tangent of 0.029 was modeled and used in the simulations as seen in Figure 157. The 0.8mm thick FR4 substrate is fixed inside the casing using some slits. There is an air gap of 1mm between the PCB and inner edge of the casing (the edge close to the head) and a 3mm air gap between the PCB and outer edge of the casing. The thickness of the dielectric casing is chosen as 1mm. The excitation of the antenna is to be done using a WFL connector (Figure 158, [18]) on the outer side of the PCB, where the CE antennas are also printed. The other face of the PCB (the face towards the head) is used as the ground plane. Three holes are left on the outer face of the casing, to get the WFL cables outside of the casing, for three antenna structures fed from different points on the PCB. This way, a generic frame and casing is obtained, to be compliant with all three antenna prototypes. This model has been manufactured using 3D printing with a semi-rigid material in the facilities of Loughborough University. The manufactured prototype can be observed in Figure 159 on a SAM head phantom.

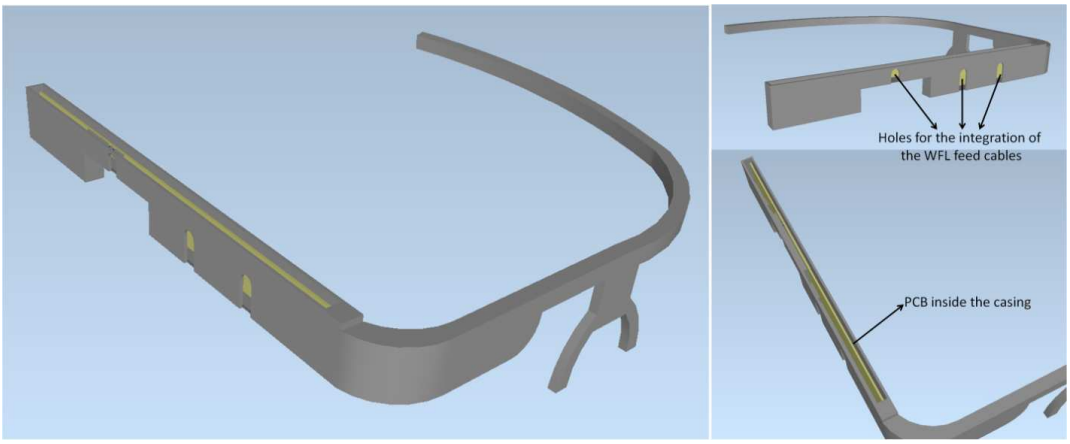

Figure 157 Eyewear Frame Model

Figure 158 WFL Connectors

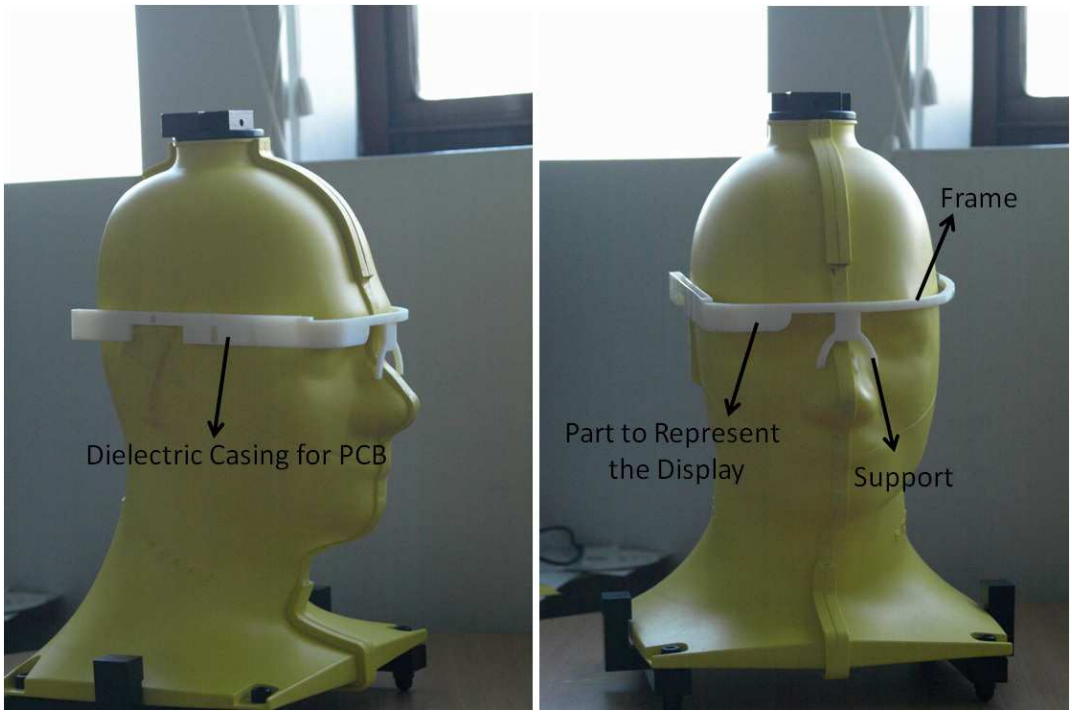

Figure 159 Manufactured Eyewear Frame

For the use cases of the eyewear antennas in real life, three scenarios were evaluated. The first scenario is the eyewear antenna and frame worn onto the head, which is foreseen as the major use case of the device. The antennas were designed and optimized for this scenario. Two other scenarios were also evaluated in the simulations, which are the free space condition and "with head+hand" condition. The free space condition can be evaluated as the use case, where the eyewear device is not worn by the user but should still be functional for the possibility of an incoming call. The other condition, "with head+hand" is the situation when the user changes the settings or the applications of the device, using the touch screen on the side with one or several fingers.

3.1. EYEWEAR ANTENNA-1

The first antenna designed is a CE, printed on the substrate, just behind the ear as seen in Figure 160. The feeding of the CE is done from the edge and a ground clearance is created at the bottom of

the CE. A matching network of 3 lumped components is introduced at the antenna feed. The simulation setups including the frame, casing, antenna, hand and the head can be observed on left of Figure 161 for the "head" configuration and on the right for the "head+hand" configuration.

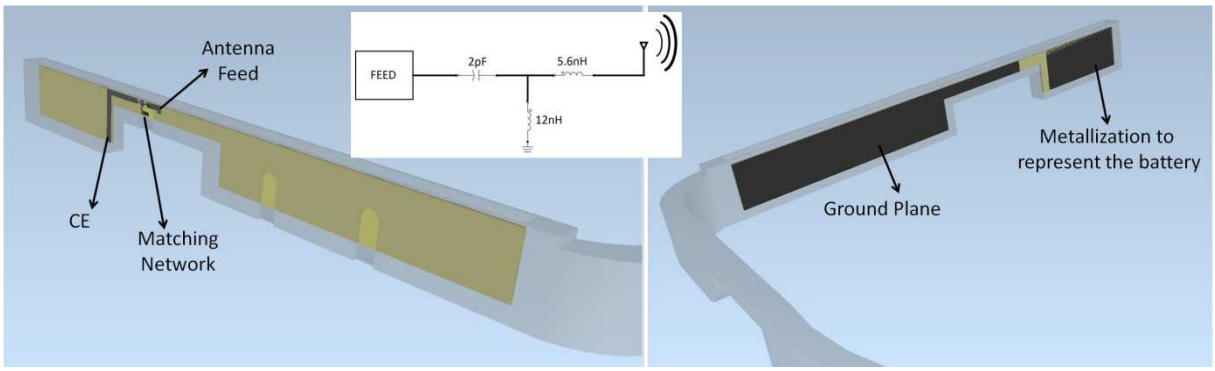

Figure 160 Eyewear-Antenna-1 Model

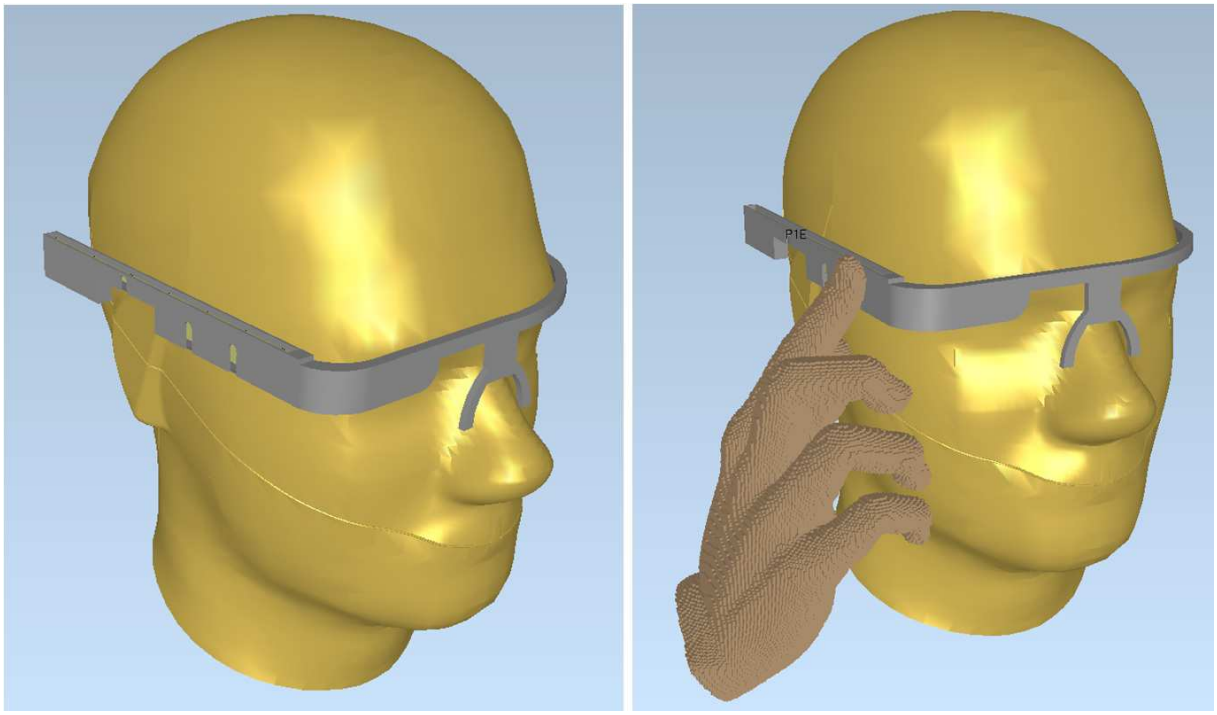

Figure 161 Positioning of the Head and Hand Phantom in Two Simulation Scenarios

The simulated input impedance, with and without the matching network can be seen in Figure 162 (in head configuration), in smith chart form plotted between 700-2700MHz. The corresponding reflection coefficient is shown in Figure 163. The antenna has a reflection coefficient below -6dB between 700-2700MHz, covering GPS operating frequency in addition to the 4G cellular communication frequency band.

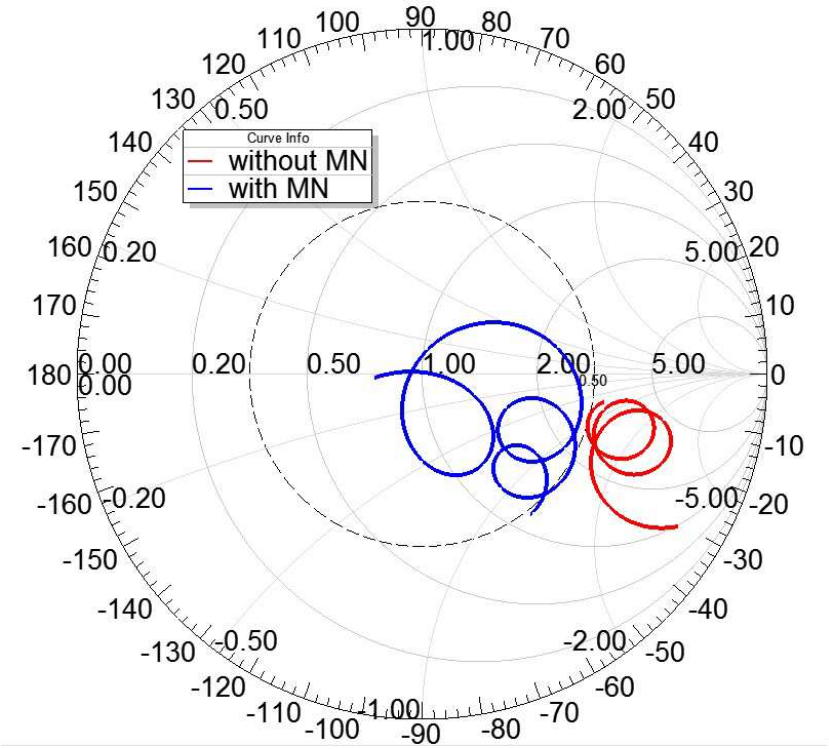

Figure 162 Simulated Impedance with Head (with and without MN) for Eyewear Antenna-1

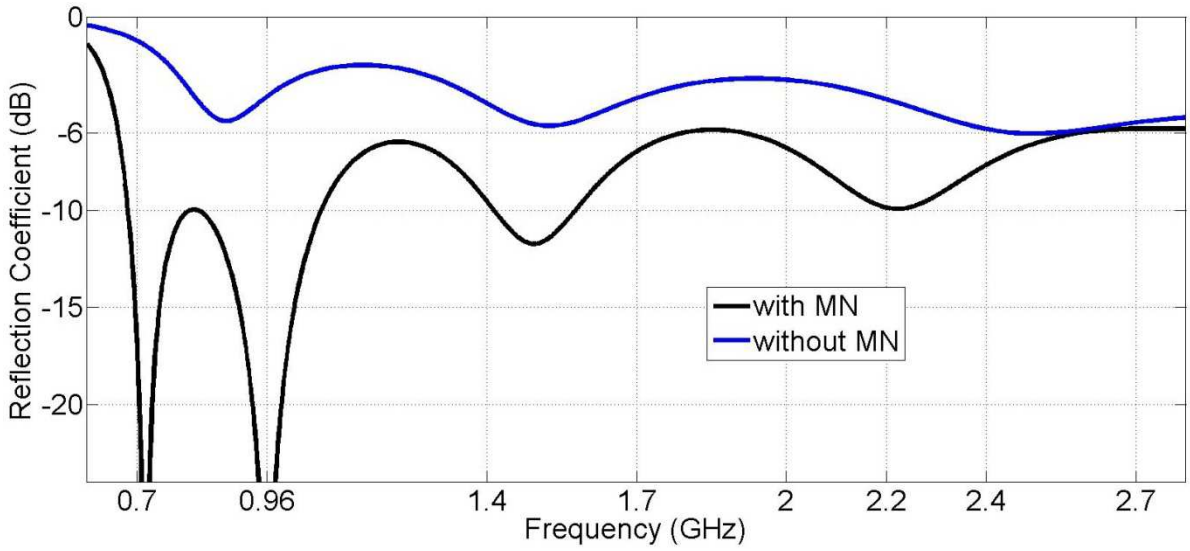

Figure 163 Simulated Reflection Coefficient (with and without MN) for Eyewear Antenna-1

Figure 164 shows the simulated 3D gain pattern of the antenna in 800Mhz and 2200MHz. The radiation is concentrated away from the head as expected due to the absorption by the head. Table 7 summarizes the simulated SAR values for three different frequencies both for 1g and 10g standards, normalized for 0.25W input power. As seen, the SAR values are high when compared to the limit in the standards (1.6W/kg for 1 gram of tissue and 2W/kg for 10 gram). The main reason is the proximity of the radiating element to the tissue in the eyewear device when compared to a typical mobile terminal, considering the antennas are generally placed on the bottom of the terminal. The SAR values can be lowered to some extent by arranging the output power from the transmitter or by increasing the distance of the antenna to the head, which will results in a thicker eyewear frame.

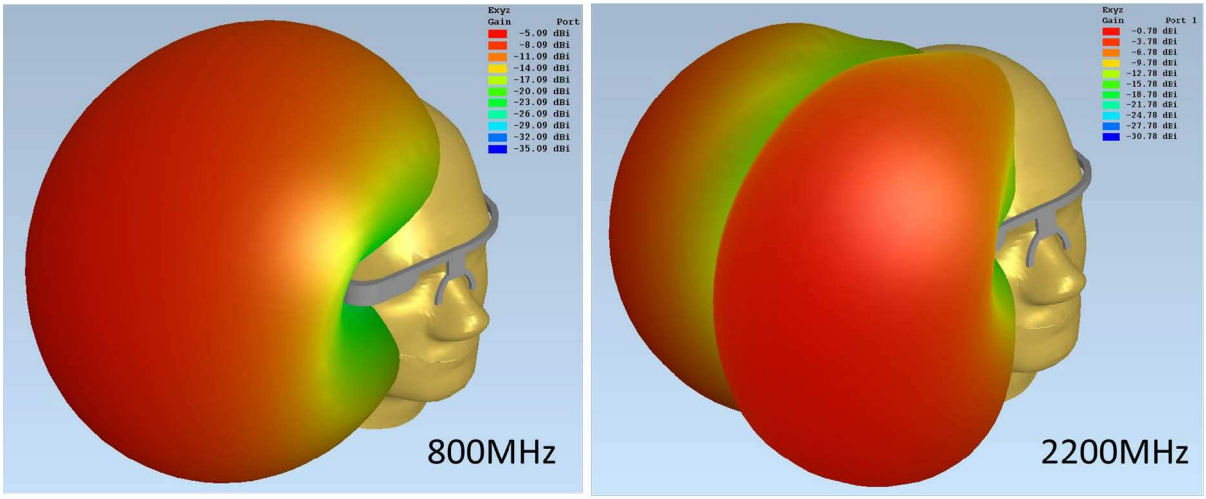

Figure 164 Simulated 3D Gain Pattern of Antenna-1

Table 7 Simulated SAR for Eyewear Antenna-1

	1g SAR	10g SAR
900MHz	5.17	2.99
2100MHz	11.28	4.90
2400MHz	8.75	3.28

The antenna was simulated also in the three use cases, mentioned before. The reflection coefficient "with head", "free space" and "with head+hand" can be observed in Figure 165. The effect of the finger close to the antenna casing do not show a significant impact on the antenna impedance and do not cause detuning, since the CE is further away from the finger. In free space, the reflection coefficient is still below -6dB in the LB, but it goes above -6dB (rising up to -3.5dB) between 1.85-2.35GHz in the HB.

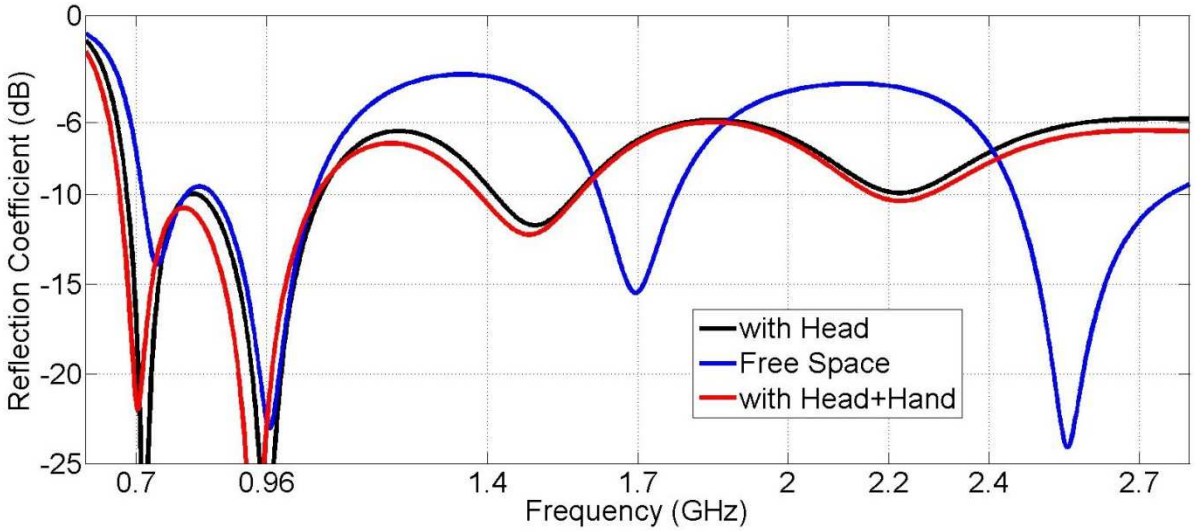

Figure 165 Simulated Reflection Coefficient for Eyewear Antenna-1 in Different Scenarios

The simulated radiation efficiency is between -13dB and -9.5dB in the LB and between -8.5dB and -6.5dB in the HB (Figure 166) in head condition. As also mentioned before, the efficiency values are considerably higher than a typical mobile terminal antenna, namely -14dB in LB and -10dB in HB for the antennas in chapter IV. The introduction of the hand do not cause significant change in the LB, but decreases the radiation efficiency by 0.5-2dB in the HB.

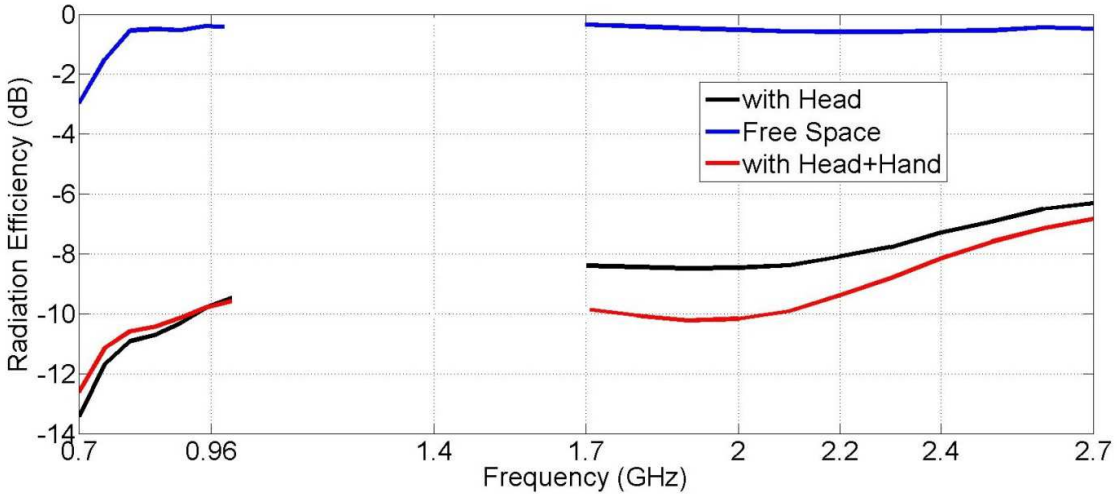

Figure 166 Simulated Radiation Efficiency for Eyewear Antenna-1 in Different Scenarios

3.2. EYEWEAR ANTENNA-2

For the second eyewear antenna design, the CE has been moved to the other edge of the PCB, close to the eye (Figure 167). There is again a MN of three lumped components at the antenna feed. The effect of the MN on the antenna input impedance can be evaluated in Figure 168, where the curves are plotted between 700-960MHz for the LB (smith chart on the left) and 1.7-2.7GHz for the HB (smith chart on right).

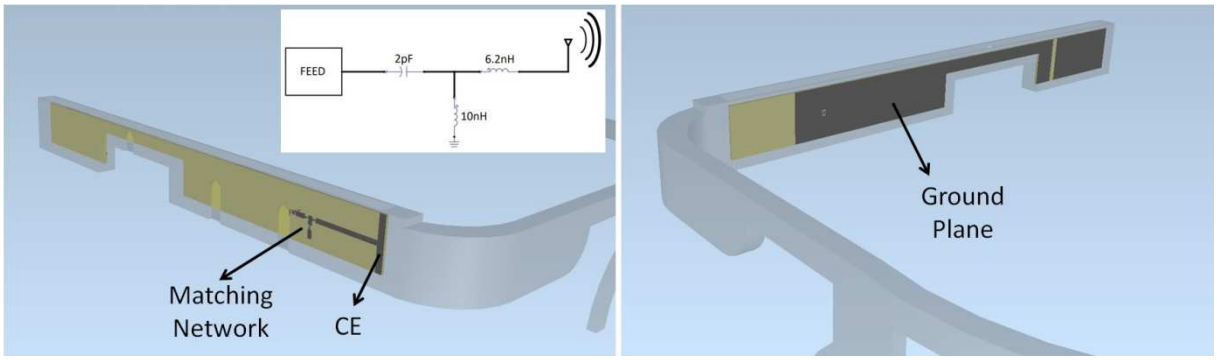

Figure 167 Eyewear-Antenna-2 Model

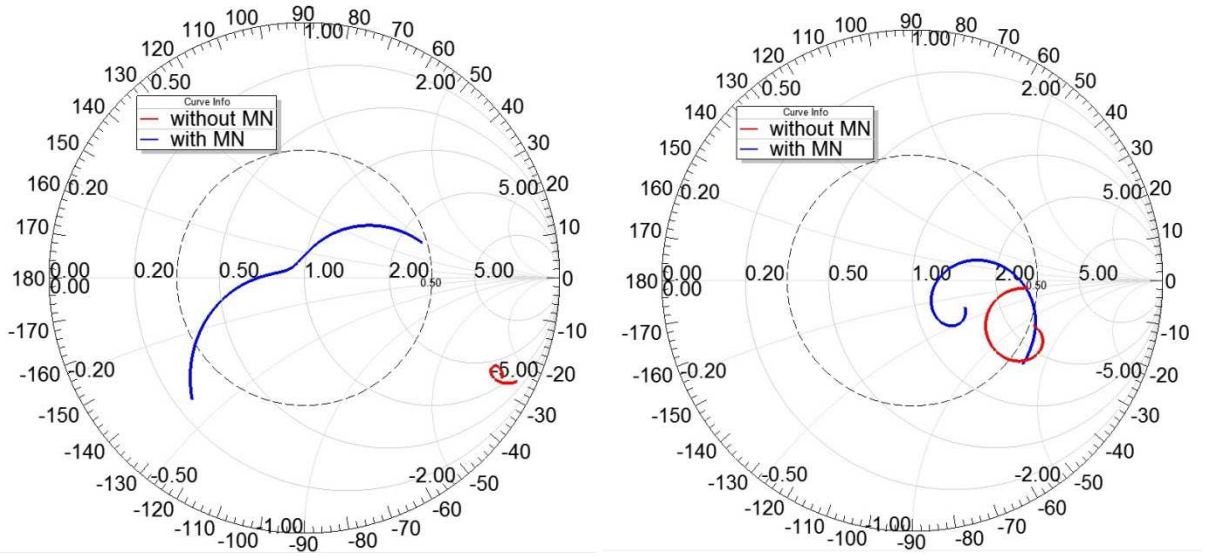

Figure 168 Simulated Impedance with Head (with and without MN) for Eyewear Antenna-2 (LB on left, HB on right)

The antenna has a reflection coefficient below -6dB between 710-960MHz in the LB and between 1.35-2.4GHz in the HB. There is slight overshoot of up to -5dB between 2.4-2.7GHz of the target band.

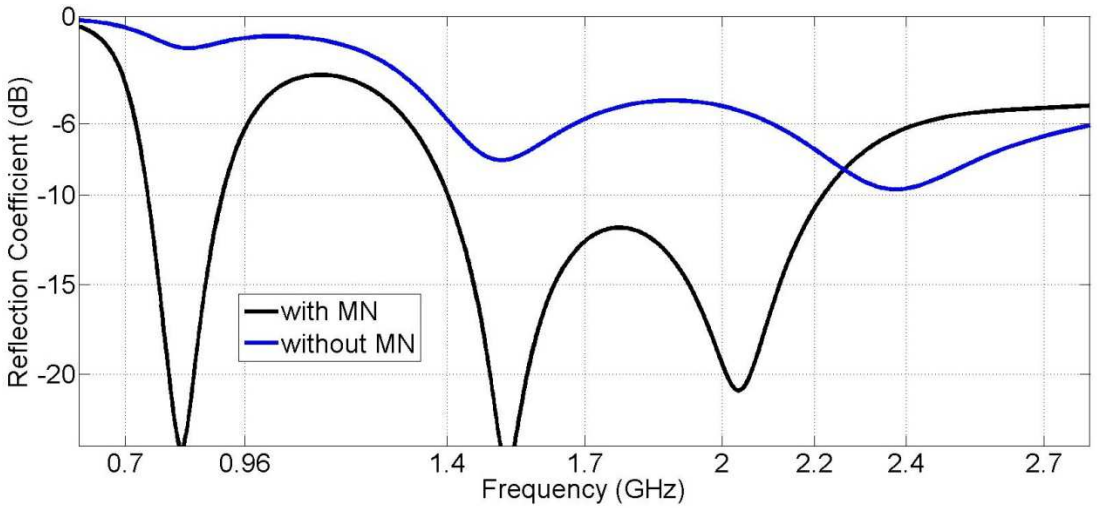

Figure 169 Simulated Reflection Coefficient (with and without MN) for Eyewear Antenna-2

Figure 170 presents the 3D gain pattern in two different frequency points. The simulated SAR values again yield high values (Table 8) as in the previous prototype, being higher than the maximum limit for 0.25W input power.

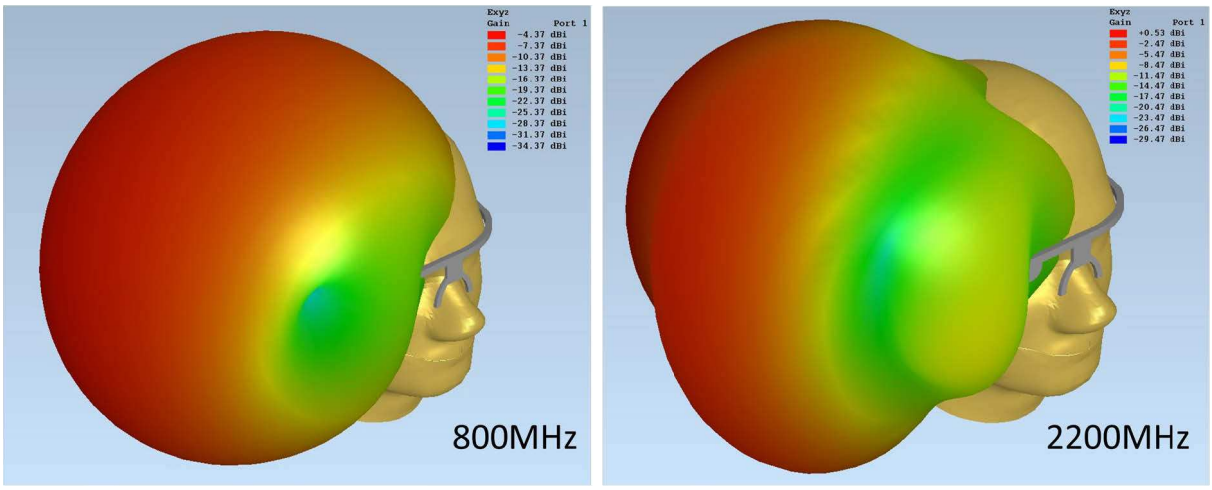

Figure 170 Simulated 3D Gain Pattern of Antenna-2

Table 8 Simulated SAR for Eyewear Antenna-2

	1g SAR	10g SAR
900MHz	4.13	2.39
2100MHz	9.80	4.62
2400MHz	7.25	3.13

When the reflection coefficient for different scenarios is observed in Figure 171, it can again be seen that the detuning is not effective due to the finger. However it is seen that the finger effect is more noticeable compared to previous antenna since the finger is closer to the radiating element. In free space, the LB performance is preserved but an overshoot up to -3.5dB is seen in HB above 2.4GHz.

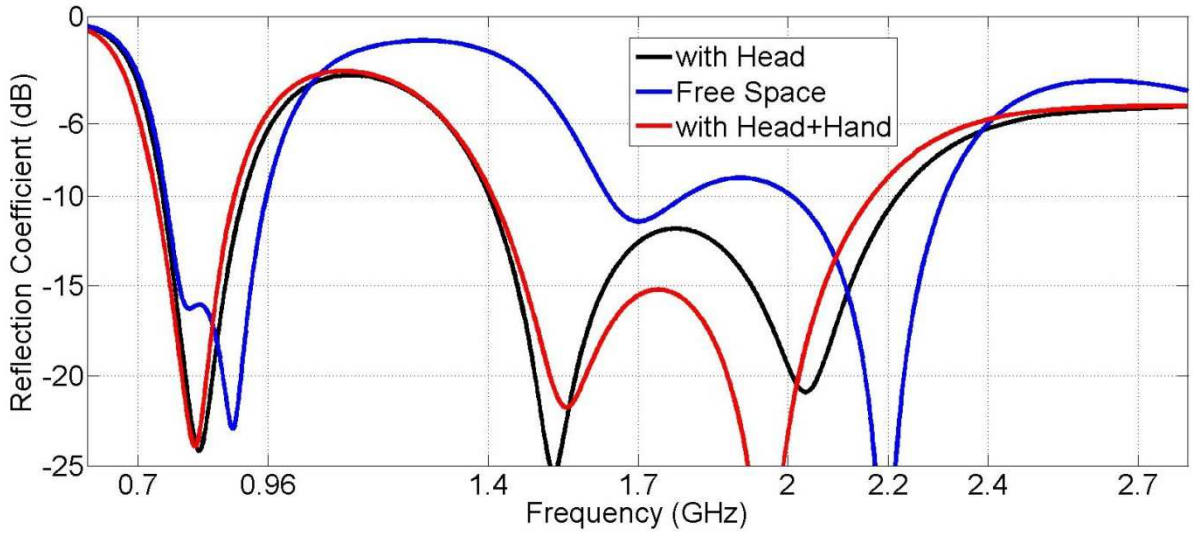

Figure 171 Simulated Reflection Coefficient for Eyewear Antenna-2 in Different Scenarios

The simulated radiation efficiency is between -11dB and -10dB in the LB and between -6.5dB and -5dB in the HB (Figure 172) for "with head" scenario. The effect on radiation efficiency is minimal

with the introduction of the hand in the LB, however a decrease of approximately 2dB is seen in the HB.

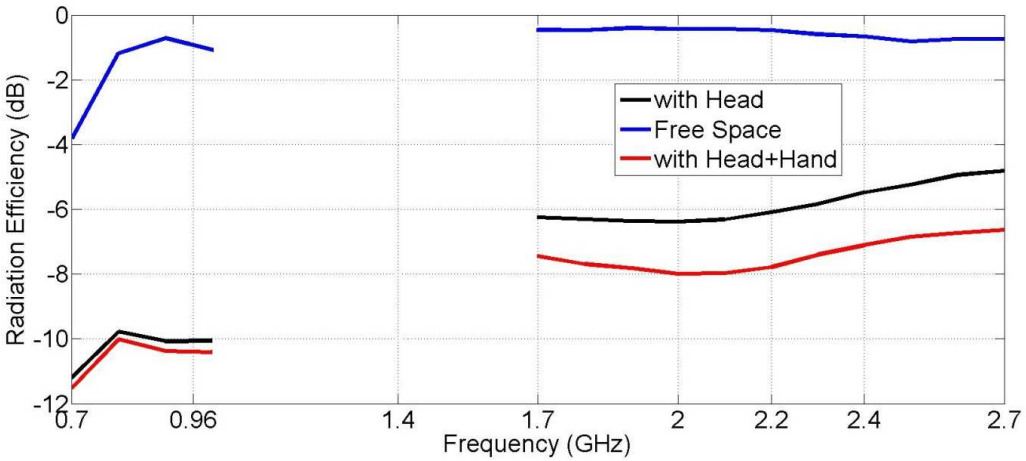

Figure 172 Simulated Radiation Efficiency for Eyewear Antenna-2 in Different Scenarios

3.3. EYEWEAR ANTENNA-3

For the third antenna design, the CE is located in the middle of the PCB section in front of the ear (Figure 173), with a MN of three lumped elements at the feed. The impedance of the antenna with and without MN is presented in Figure 174.

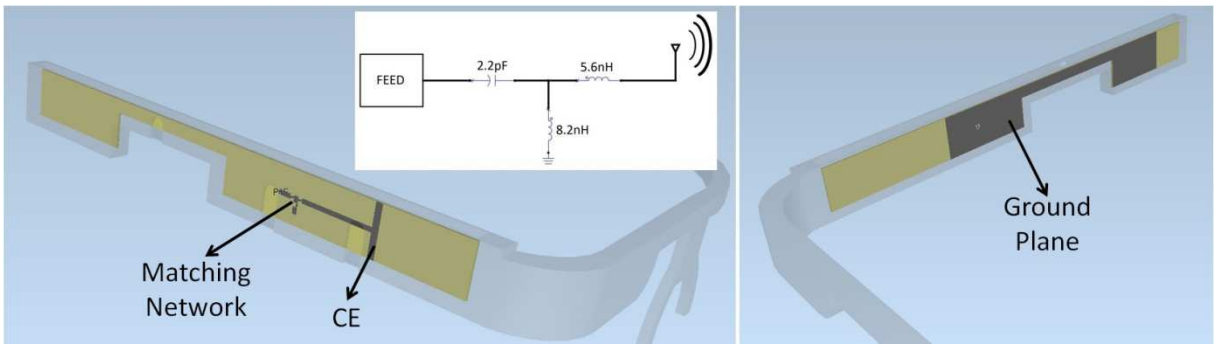

Figure 173 Eyewear-Antenna-3 Model

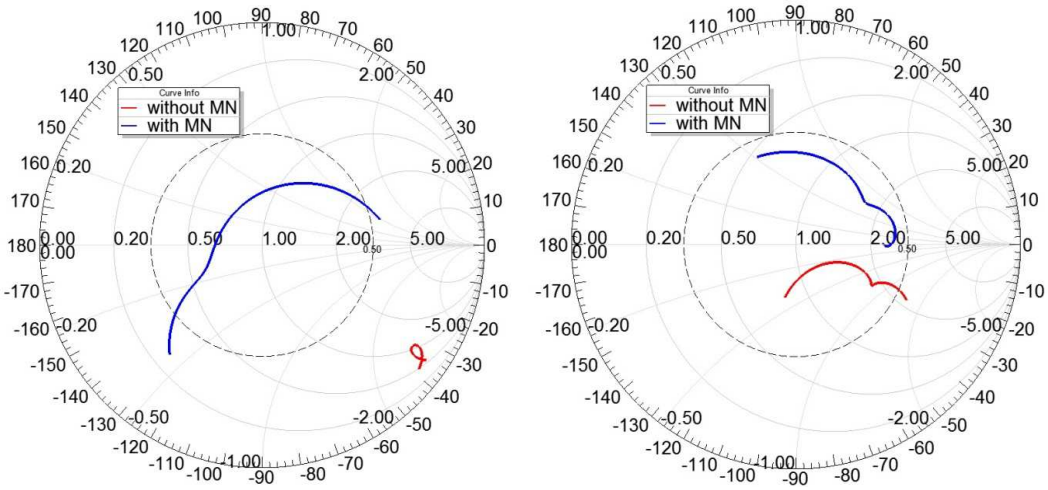

Figure 174 Simulated Impedance with Head (with and without MN) for Eyewear Antenna-3 (LB on left, HB on right)

The third antenna proposed, has a reflection below -6dB between 710-950MHz and 1.55-2.7GHz (Figure 175).

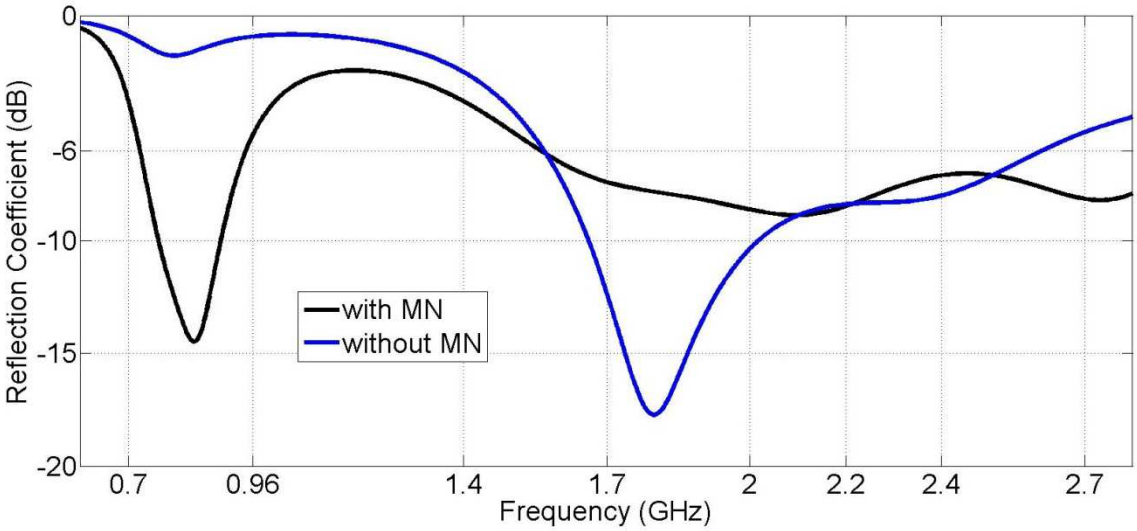

Figure 175 Simulated Reflection Coefficient (with and without MN) for Eyewear Antenna-3

Figure 176 shows the simulated 3D pattern of the antenna. The results of the SAR simulations for three frequency points are given in Table 9.

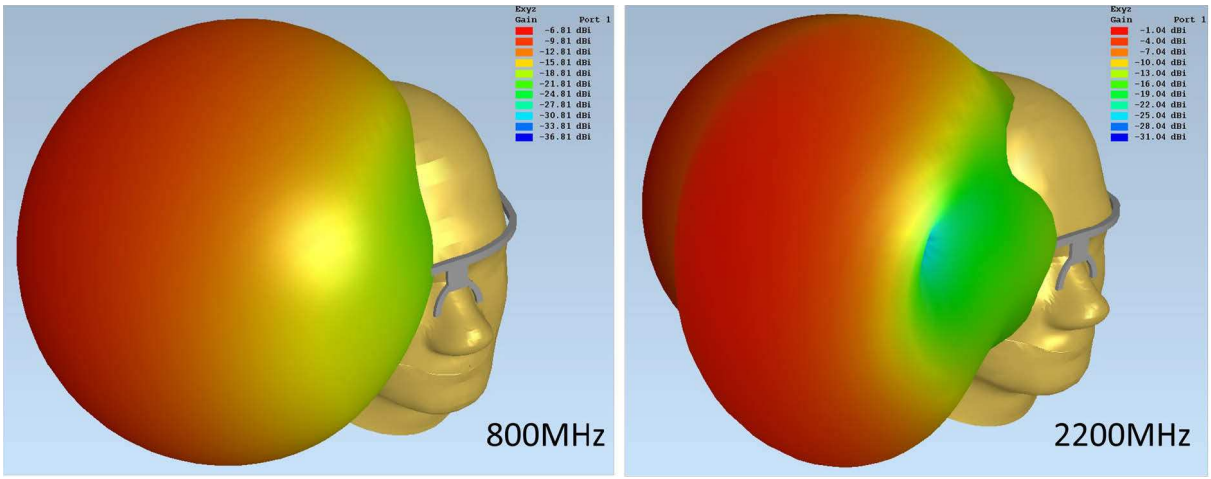

Figure 176 Simulated 3D Gain Pattern of Antenna-3

Table 9 Simulated SAR for Eyewear Antenna-3

	1g SAR	10g SAR
900MHz	4.67	2.55
2100MHz	8.77	4.18
2400MHz	7.76	3.51

As in previous designs, the hand does not cause any detuning on the antenna reflection coefficient(Figure 177). However, in free space condition, the HB coverage is effect significantly, having a reflection coefficient below -6dB only between 2-2.55GHz.

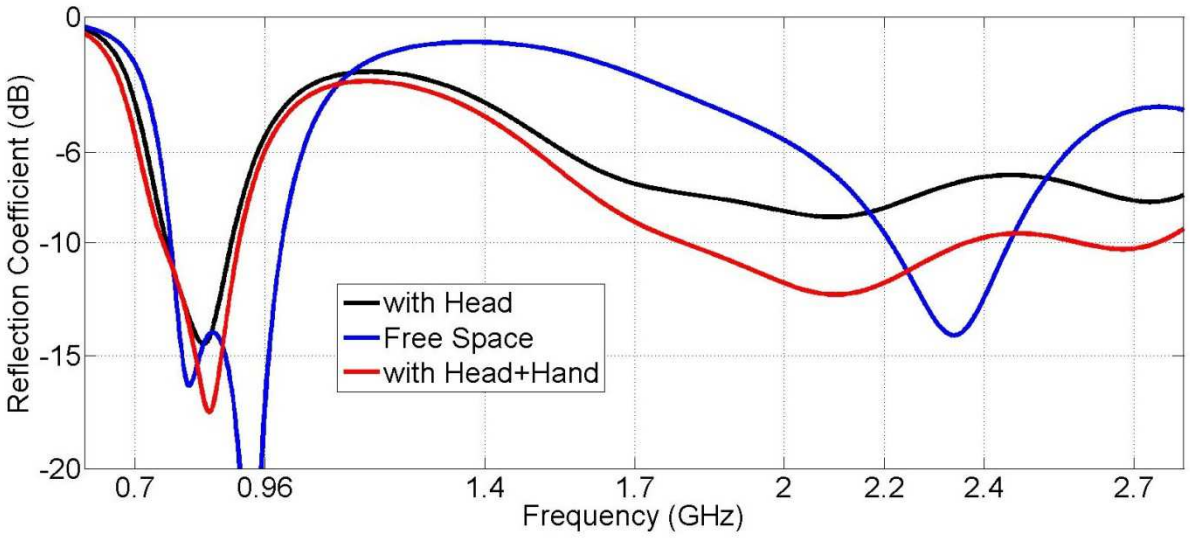

Figure 177 Simulated Reflection Coefficient for Eyewear Antenna-3 in Different Scenarios

The simulated radiation efficiency is between -14dB and -12.5dB in the LB and between -8dB and -6.5dB in the HB (Figure 178) with the head. The hand causes an efficiency drop of up to 1.5dB in the HB.

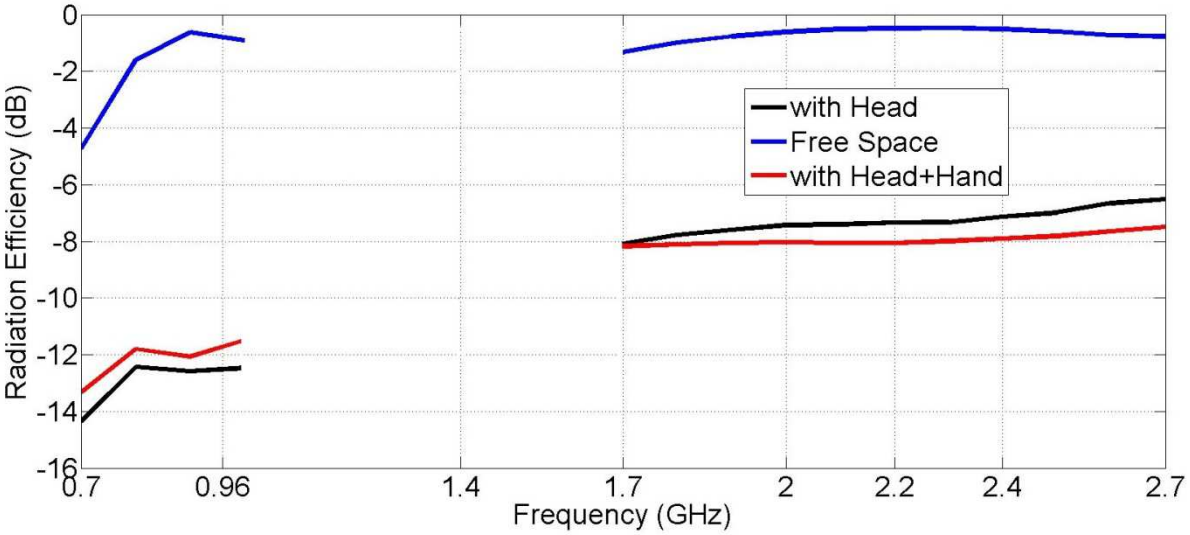

Figure 178 Simulated Radiation Efficiency for Eyewear Antenna-3 in Different Scenarios

4. CONCLUSION

In this chapter, an application example of CEs was presented, differently from mobile terminal antennas. The CEs were used to obtain coverage in the 4G cellular communication bands, in an eyewear structure, which are gaining popularity in the market as wirelessly connected wearable devices. Firstly, a feasibility study was performed using three different locations (behind the ear, in front of the ear and close to the eye) for the CEs, comparing the bandwidth potential and the expected SAR value. It was shown through the simulation results that placing the antenna just in front of the ear was giving the worst bandwidth potential. The potential was higher in "close to the eye" location and the highest behind the ear. It should be noted that these results are subject to

change due to modifications on the system ground plane. For the SAR, the minimum value was observed with "close to the eye" location since the radiating structure is further from the head when compared to the other two locations.

Taking into account the feasibility study, three antennas were designed, with a dielectric casing & frame. The first antenna, which is placed behind the ear, had the highest operating bandwidth with a continuous coverage from 700MHz up to 2.7GHz with a three element MN at the feed. The other two designs cover the 700-960MHz and 1.7-2.7GHz. The simulated radiation efficiencies are generally on the level of -12dB for the LB and -7dB for the HB. However, there is a significant increase in the radiation efficiency when the CE is placed close to the eye, since the head is further from the antenna, reducing the losses. For evaluating the real use case of the device, simulations were performed also with the user hand and in free space condition. It was shown through simulations that the hand does not cause a significant detuning, however poorer reflection coefficient behavior may be seen in free space, as expected.

5. PERSPECTIVES

The first target concerning the eyewear antennas is the fabrication of the antennas and the corresponding PCBs with MNs. Using the PCBs with the already manufactured dielectric frames, the measurements for the S-Parameters and total efficiency will be realized considering free space, with head and with hand+head conditions, as well as SAR measurements.

For the future steps, a MIMO scenario will be considered. The first idea about obtaining MIMO operation is to place one antenna on the right side of the head and the other to the left side. In this way, a good isolation level was observed in the initial simulations since the head acts as an absorbing element between the antennas, increasing the isolation. Considering the general characteristics of the radiation pattern (radiating away from the head), a low ECC value is also foreseen with this topology.

Another possible investigation will be based on the effects of the metallic frames instead of dielectric frames. The effect of metallic frames on the SAR of mobile terminals has been previously investigated in [19-23], showing the possibility of significant effects. For this reason, simulation and measurements will be repeated with metallic frames to compare the general antenna characteristics (with dielectric frame) as well as SAR levels.

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CONCLUSION

Several different antenna designs have been proposed in this thesis, for mobile terminals operating in the 4G frequency bands, i.e 700-960MHz and 1.7-2.7GHz. A variety of strategies have been evaluated like single/multi-feed and also passive/active antennas. Antenna alternatives for different mobile communication devices (like electronic eyewear) operating in 4G standards have also been mentioned.

Tunable components have been evaluated in circuits placed at the feed of mobile terminal antennas to achieve frequency band reconfigurability in Chapter II. MEMS switches and BST tunable capacitors were used as the tunable components, which are under development in the scope of the Artemos project by partners. Two antennas exploiting the MEMS switches were proposed and the main aim was to obtain a band switching behavior, to select either LB (700-960MHz) or HB (1.7-2.7GHz) operation. Two such antenna designs were proposed, where one of the design uses two SPDT MEMS switches to select between one of the two MN branches and the other uses only one switch to select one of the two driven strips. Due to components' reliability difficulties, S-Parameter measurements could be performed for only one of these prototypes. For the measurement of the other prototype, passive antennas were manufactured and band-switching performance was verified this way. The target bands could be covered comfortably with a reflection coefficient below -6dB (observed as generally below -7dB or -8dB) in the measurements. The passive prototypes were also used for the total efficiency measurements, which are on average 50% in the LB and 65% in the HB. It should be noted that these values will for sure be subject to a slight decrease taking into account the insertion losses of the MEMS switches. Another antenna design was also proposed using a BST capacitor to tune the input impedance of the antenna in the 700-960MHz frequency range. The coverage of the HB was achieved with a separately fed antenna composing a dual-feed system. The measurement results of this antenna is planned to be done as soon as the integration of the BST capacitor to the antenna PCB is done.

The main advantage of the antennas with tunable components is the relatively lower reflection coefficient obtained especially in the LB (around -8dB), when compared to other passive prototypes. This would result in higher total efficiency than passive prototypes however considering also the insertion losses of the tunable elements and also the effect of the bias lines, a similar level in terms of total efficiency can be foreseen. On the other hand, the main drawback was the reliability of the tunable elements. The main reason is the continuing design & optimization phase for these components by their manufacturers (in Artemos project). More reliability and repeatability is foreseen after the finalization of the optimization phase of the components. Another disadvantage when compared to passive prototypes is the additional losses due to internal resistances of the tunable elements, which will in return cause decrease in the total efficiency. There is also the disadvantage about the carrier aggregation for LTE-A standards. For example for the band-switching topologies (with MEMS switches), it is not possible to use inter-band carrier aggregation since either the LB or HB is matched at a time.

In Chapter II, a study about the circuit model representation of a driven strip with a grounded parasitic strip was also carried out. The accordance of the circuit model was verified through electromagnetic simulations. Another issue addressed in this chapter was the radiation efficiency drop, that occurs in the HB for the antennas using parasitic strips. Solution techniques have been proposed using some modified topologies.

A summary of the tunable antennas mentioned in Chapter II can be seen in Table 10. The green shaded results in the table indicate measurement results with passive prototypes, blue shaded results indicate active measurements and yellow fields indicate simulations.

Table 10 Summary of the Performance of Tunable Antennas

	S11(dB) in LB		S11(dB) in HB		Eff(dB) in LB		Eff(dB) in HB		PCB Length	GND Length	Height
	Max	Min	Max	Min	Min	Max	Min	Max			
MEMS-1	-7	-16.2	-3.8	-30	- 3.7	-1.5	-3.5	-1.1	120mm	107mm	-
MEMS-2	-5.7	-17	-11.3	-33.7	-4.9	-1.7	-2.5	-1.1	115mm	100mm	-
BST	-3.2	-27.3	-4.9	-32	-	-	-	-	120mm	105mm	-

Dual-feed antenna structures using coupling elements were proposed in Chapter III, again covering the 4G standard frequencies. In these designs, the plain CE which was used extensively in the literature for mobile terminal antennas was modified, making it hollow. The coverage of the LB was achieved using these hollow CEs, with appropriate matching networks. Separately-fed monopole type antennas were used for the HB coverage, obtaining high isolation levels thanks to both antenna placement and band-stop characteristics of the matching networks. Measured total efficiency levels of 40-50% and 50% (on average) were obtained in the LB and HB respectively, which are over the 50% level generally expected in free space for mobile terminal antennas. These antennas had the advantage of being appropriate for both intra-band and inter-band carrier aggregation. A further advantage of the dual-feed topology is the opportunity to get rid of a diplexer in the RF front-end for inter-band carrier aggregation, which is definitely needed in a single feed antenna. An application example of the hollow CEs was further proposed in the same chapter, by placing two CEs on the corners of the PCB, to achieve MIMO operation in the LB (between 740-960MHz). The isolation between the two elements in lower LTE band was enhanced using a neutralization line between the antennas.

In Chapter IV, single-feed antenna designs manufactured using LDS technology were presented. Both antennas consist of a driven strip which capacitively excites a grounded parasitic strip for bandwidth enhancement. One of these antennas covers the low LTE/GSM/UMTS bands, whereas the other covers all the 4G bands. The main challenge for these antennas was the short ground plane (75mm, considered because of the 3.5 inch touchscreens common in most of modern smartphones), which makes it difficult to cover down to 700MHz. The total efficiencies of the antennas with 75mm ground plane were on average 35% in LB and 50% in HB. When the ground plane length was increased to 90mm, the LB efficiency was increased by approximately 10%, without any significant change in the HB. The effect of the user's hand and head was investigated on these antennas through measurements, to form a basis for future studies. The first results of this study were also presented at the end of the chapter, proposing an antenna topology resistant to detuning due to the hand of the user (in simulations only).

A comparison of passive antenna prototypes from Chapter III and Chapter IV is presented in Table 11, in terms of dimensions, reflection coefficient and efficiency. All the values presented in the table are measured results.

Table 11 Summary of the Performance of Passive Antennas

	S11(dB) in LB		S11(dB) in HB		Eff(dB) in LB		Eff(dB) in HB		PCB Length	GND Length	Height
	Max	Min	Max	Min	Min	Max	Min	Max			
HollowCE	-4.7	-9.1	-2.8	-27.5	-6	-1.9	-4.3	-1	114mm	100mm	3mm
PrintedCE	-3.1	-9	-3.8	-15	-5.1	-1.5	-4.3	-1.2	114mm	100mm	-
UMTS LDS	-5	-22	-5.2	-19	-7.7	-3	-3.9	-2	114mm	75mm	5mm
LTE LDS	-3.8	-15.9	-3.5	-16.3	-14	-3.1	-5.5	-1.7	114mm	75mm	5mm

Finally in Chapter V, antennas using coupling elements were proposed for 4G operation in an eyewear antenna. First, a feasibility study was presented by evaluating three different positions for the CE on the PCB of the electronic eyewear device, considering the bandwidth potential and SAR. Afterwards, three antennas were proposed with appropriate matching networks. The behavior of these three antennas was evaluated in different use cases like free-space, with the head and with head and hand (to stand for the user changing applications or settings by putting his finger on the touchpad).

To sum up, it can be told that the desired performances in terms of reflection coefficient (below -6dB) and total efficiency (higher than 50% in free space) have been achieved with the novel antenna designs in this thesis. A final note can be made be on the utilization of active tunable elements, comparing the tunable/passive antenna designs in this thesis. As presented, the passive antennas can reach the target performance without any tunable components in free space condition. This way, the additional losses, complexity (like biasing circuit) and reliability concerns from the tunable elements can be avoided. However, the utilization of the tunable components can be very useful for minimizing the effect of the user and increasing the total efficiency of the antenna for instance in a tunable matching unit at the antenna feed .

FUTURE PERSPECTIVES

A possible improvement for the antennas presented in this thesis can be to minimize the effect of the user on the reflection coefficient and by this way increase the total efficiency taking into account the user's hand, head or both. The user effect (hand and head) have been investigated for the single-feed antennas in Chapter IV (antennas on LDS) and it was seen through measurements that the antennas (placed on the bottom of PCB) get detuned (towards lower frequencies) when the hand of the user is taken into account.

A different antenna design technique was recently proposed in [1-3], considering in the design phase the matching performance of the antenna in the presence of the user. This technique relies on combining the antenna design with circuit design concepts to cover 700-960MHz, 1.7-2.2GHz and 3.3-3.8GHz frequency bands [1]. The novelty of the design is to transfer the complexity of the antenna design to the complexity of the circuit before the antenna (at the antenna feed). The main aim is to get a very wideband antenna so the detuning effect due to the user (which unavoidably exists) is minimized by this way. In these designs, multiple simple IFA type radiators are used. The first and second order resonances of these radiators are selected and combined together to get a wideband response using diplexers and power divider/combiners resulting in a single-feed topology. These circuitries are composed of lumped SMD inductors/capacitors with up to 27 lumped elements.

This approach proves to be very robust to detuning due to user effects in cases like right/left hand and with head (in right/left hand configurations). When this methodology is compared with the results of the antennas in Chapter IV of this thesis, it is evident that the total efficiency is considerably lower (on average) in this methodology for free space condition (between 5-40% in LB and 15-55% in HB). This relatively low efficiency may cause a problem for the mobile phone manufacturers during the manufacturing tests in free space condition. Although robust to detuning, the measured efficiency with the user hand in [1] (5% in LB, 10% in HB on average) is also lower by approximately 3dB when compared to the antennas in this thesis. The results are approximately the same for the measurement with the head. However, one good advantage of this approach is that the designs can be integrated to any mobile phones by changing the values of the components in the circuitry.

Keeping this in mind, the main focus for the future studies on the user effects of the antennas in this thesis should be to minimize the detuning. For this purpose, some rule of thumbs were deduced in Chapter IV, from the measurements with hand/head to foresee the possible changes in the input impedance with user effects. Using this background, the first results from a dual-feed antenna design was presented with both simulations and measurements.

Another study was also started to design a wideband antenna as to cover 700MHz-2700MHz continuously, both to minimize the detuning due to user effects and also to enable the coverage of possible other standards (like GPS). The main target is to obtain a very high bandwidth potential around 1700MHz (which is the center frequency in this case) and thus obtain a very wideband antenna. For this purpose, the first simulation results of the antenna given in Figure 179 can be seen in Figure 180. The antenna consists of a driven strip printed under the plastic piece (same as the plastic pieces in antennas manufactured with LDS technology), which capacitively excites a loop antenna connected to the system ground plane at both ends. To increase the electrical length of the loop, two lumped inductors (each 25nH) are modeled in series with the loop, located at the

grounding points. A two element matching network consisting of a shunt capacitor (0.6pF) and series inductor (5.6nH) is used at the feed.

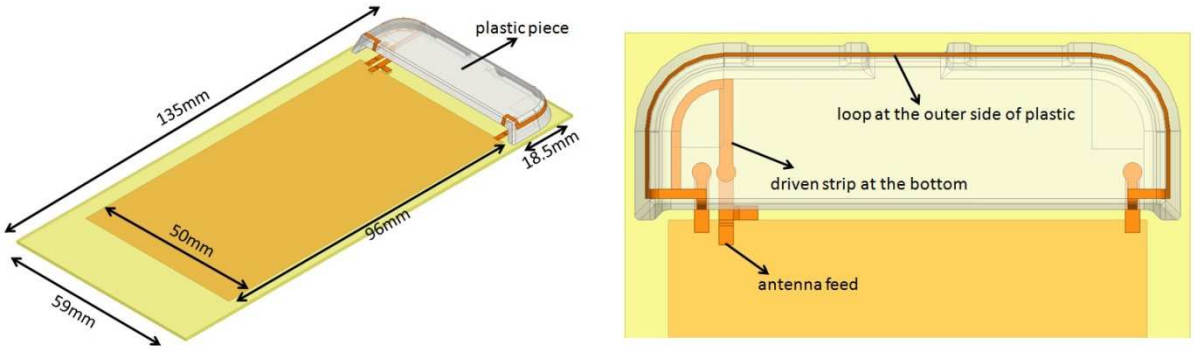


Figure 179 Parasitic Loop Antenna Model

As seen in Figure 180, the antenna can cover the interval of 700-2700MHz continuously with a reflection coefficient below -6dB. The simulated radiation efficiency of the antenna is given in Figure 181. A relatively low radiation efficiency is seen around 700MHz (30%) whose main reason is the high loss resistance values of the 25nH inductors modeled in simulations. It is foreseen that the efficiency can be increased by using lower value inductors (hence lower losses), by increasing the physical length of the loop by meandering.

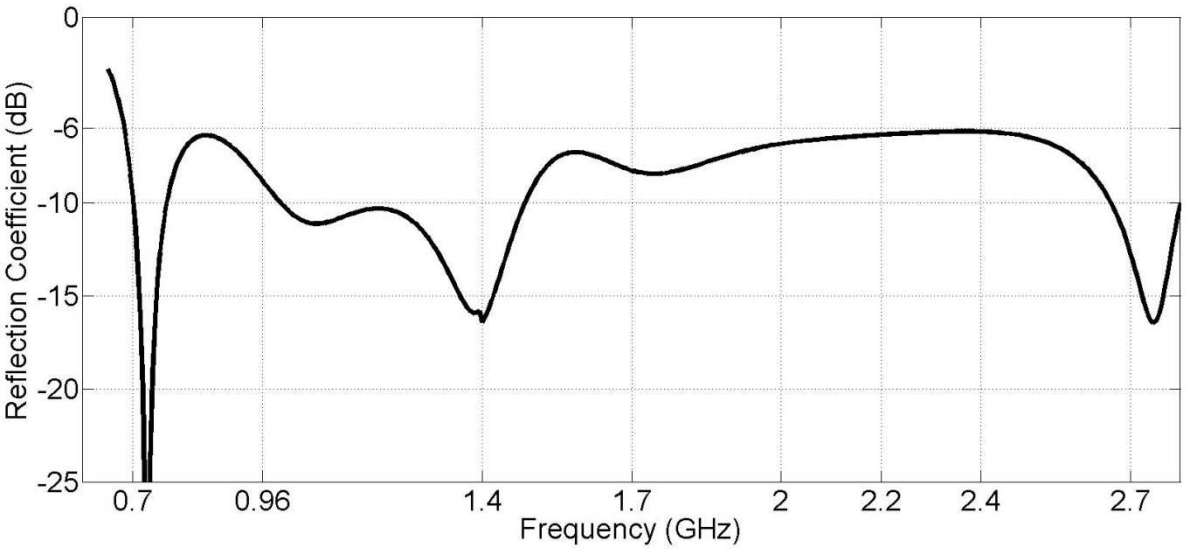
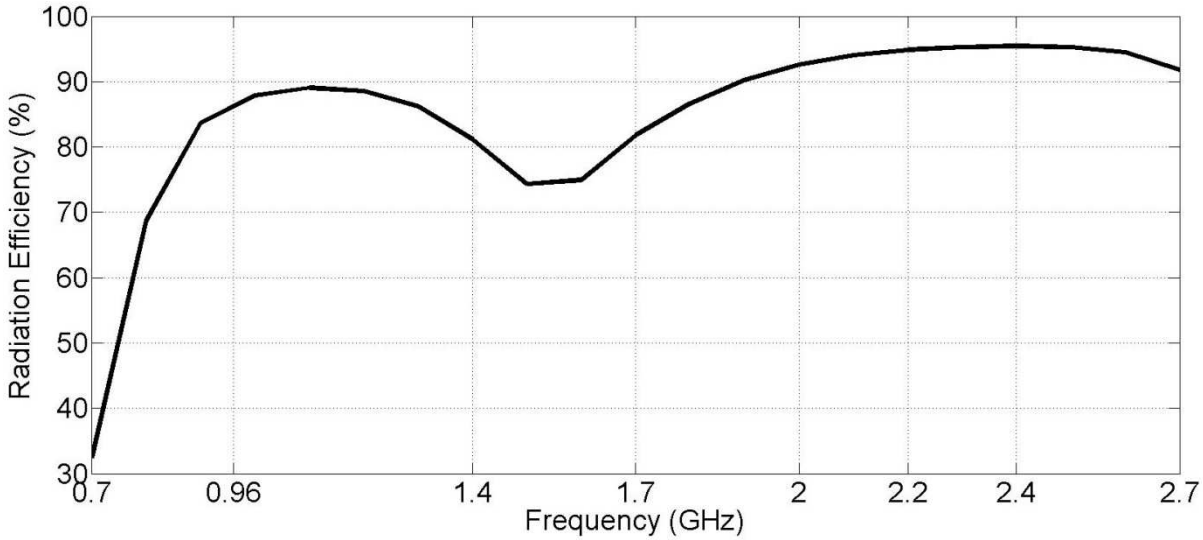


Figure 180 Simulated Reflection Coefficient of the Parasitic Loop Antenna



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1. **A. Cihangir**, F. Sonnerat, F. Ferrero, R. Pilard, F. Gianesello, D. Gloria, P. Brachat, G. Jacquemod, C. Luxey, "Neutralisation technique applied to two coupling element antennas to cover low LTE and GSM communication standards," Electronics Letters , vol.49, no.13, pp.781,782, June 20 2013

2. **A. Cihangir**, W. Whittow, C. Panagamuwa, F. Ferrero, G. Jacquemod, F. Gianesello, C. Luxey, "Feasibility study of 4G cellular antennas for eyewear communicating devices," Antennas and Wireless Propagation Letters [ACCEPTED FOR PUBLICATION]

3. **A. Cihangir**, F. Ferrero, G. Jacquemod, P. Brachat, C. Luxey, "Neutralized Coupling Elements for MIMO Operation in 4G Mobile Terminals," Antennas and Wireless Propagation Letters [ACCEPTED FOR PUBLICATION]

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2. **A Cihangir**, F. Sonnerat, F. Ferrero, C. Luxey, R. Pilard, F. Gianesello, G. Jacquemod, "Design of traditional and a novel space-efficient antenna-coupling elements for lower LTE/GSM mobile phones," Loughborough Antennas and Propagation Conference (LAPC 2012)

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1. **A. Cihangir**, F. Ferrero, C. Luxey, G. Jacquemod, P. Brachat, "Conception d'un nouvel élément antennaire de couplage pour terminaux mobiles LTE/UMTS/GSM," Journées Nationales Microondes (JNM), 2013

2. **A. Cihangir**, F. SONNERAT, F. Ferrero, R. Pilard, F. Gianesello, D. Gloria, C. Luxey, G. Jacquemod, P. Brachat, "Antenne en technologie LDS pour terminaux mobiles LTE700 et GSM850/900," Journées Nationales Microondes (JNM), 2013

AWARDS

3rd place in "IET Best Student Paper" contest in LAPC2013

ABSTRACT

With the introduction of the fourth generation (4G) of cellular mobile communications, there has been an increase in the frequency bands that a generic mobile phone should be able to operate. This requirement poses some challenges in the design of the transceiver modules of the device as well as the RF front-end and the antenna.

From the antenna design point of view, the newly added frequency bands especially LTE bands 12-13-17 which goes down to 700MHz creates the main design challenge: covering 700-960MHz (30% bandwidth) in the low-band, considering also backward compatibility with the former technologies like GSM850/900. Since the space reserved for the antenna is electrically small in this frequency range, special design techniques are necessary to obtain the required bandwidth for a generic mobile terminal antenna.

It should also be taken into account in the design phase, that the antenna needs to be appropriate for operation in the data-rate enhancement techniques like MIMO (multiple input multiple output) and carrier aggregation in LTE-Advanced. Furthermore, considering the real use cases of the mobile terminal, the effect of the user (hand, head and both) on the antenna performance should also be investigated.

In this thesis, several antenna designs for 4G coverage in mobile terminals are proposed. Different antenna topologies are evaluated consisting of tunable/passive antennas as well as single/multi-feed structures.

RÉSUMÉ

Avec l'introduction des standards de communication de quatrième génération (4G) dans la téléphonie mobile, les bandes de fréquence dans lesquelles un téléphone mobile générique doit fonctionner ont naturellement augmenté. Ces évolutions ont induit de nouveaux challenges techniques notamment en ce qui concerne la conception des modules d'émission-réception, du frontal RF mais surtout des antennes.

Les bandes de fréquences nouvellement ajoutées, en particulier les bandes LTE 12/13/17 qui descendent jusqu'à 700 MHz, constituent le défi antennaire majeur: couvrir la bande basse des 700-960MHz (30%) notamment en incluant la bande GSM850/900 compte tenu de la rétrocompatibilité avec les standards 3G. Etant donné que l'espace réservé pour l'antenne dans un terminal portatif est électriquement petit (dans cette gamme de fréquences basses), des techniques de conception spécifiques doivent être mises en œuvre pour pouvoir couvrir toute la bande de fréquence basse.

Dans la phase de conception, il doit également être pris en compte le fait que l'antenne ou les antennes puissent incorporer des techniques d'amélioration de débit de données telles que le MIMO (Multiple Input Multiple Output) et l'agrégation de porteuse (Carrier Aggregation) pour le LTE-Advanced. De plus, si l'on considère des cas d'utilisation réalistes, les effets de l'utilisateur (main, tête) sur les performances globales du terminal doivent aussi être examinés.

Dans cette thèse, plusieurs concepts d'antennes 4G pour les futurs terminaux mobiles sont proposés. Différentes topologies antennaires sont évaluées notamment des antennes reconfigurables / passives ainsi que des structures à alimentation simples et multiples.

1 Monitoring for cloud apps AI powered, full stack, automated — Free trial! dynatrace.com

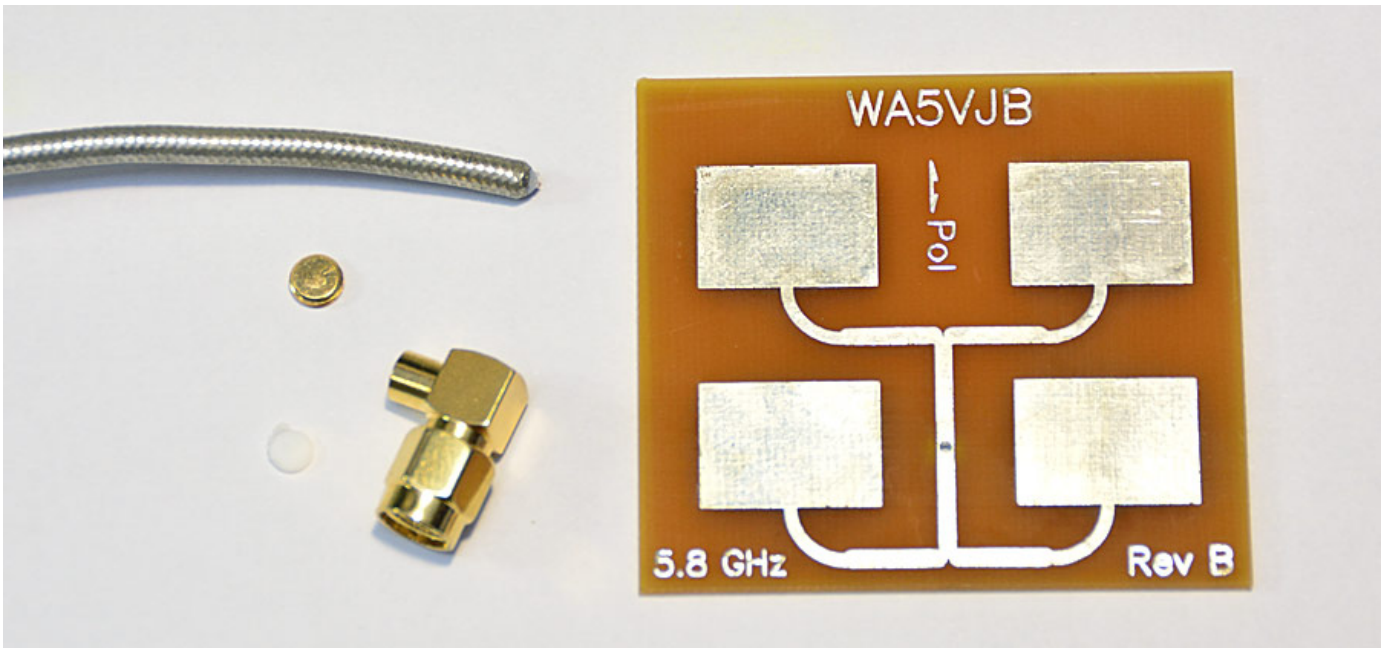
2 Voted #1 VPN Service Full Access to Censored Apps & Websites. 100% Secure & Anonymous. Fastest Speeds. ipvanish.com

Homebrew 5.8 GHz patch FPV antenna

Posted on September 26, 2014 (http://vk4ghz.com/homebrew-5-8-ghz-patch-fpv-antenna/) Posted in FPV (http://vk4ghz.com/category/multirotors/fpv/), Multirotors (http://vk4ghz.com/category/multirotors/)

Like Share 35 people like this. Be the first of your friends.

These 5.8 GHz PCB patch antennas come from Kent Electronics (http://www.wa5vjb.com/) in Texas, USA.

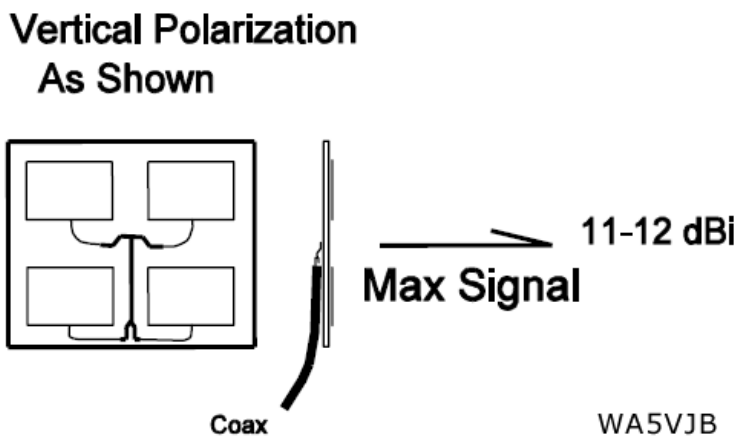


You can download the product PDF from here (http://www.wa5vjb.com/pcb-pdfs/5800MHzPatchArrays.pdf).

The quad patch array (pictured above) offers 11 – 12dBi gain, or 8.8 – 9.8 dB gain over a (unity gain) dipole.

To put that into a simplistic terms, is almost 10 times better than your dipole.

Along with some UT141 semi-rigid hardline coaxial cable, and a right-angle SMA (m) connector, both from RF Supplier (http://www.rfsupplier.com/) in Shenzhen, China, a low-cost patch antenna can be made for multirotor FPV use.



Another handy store for connectors and coax in Brisbane is The RF Shop (http://rfshop.com.au/), located in Oxley.
RF Shop is more expensive than a Chinese online store, (and most of their product is better quality, from Taiwan), and operates out of a home, but give them a call first, as they don't mind if you drop in and pay & pick up.

First step, prepare one end of the 141 to be soldered to the SMA (m) connector.

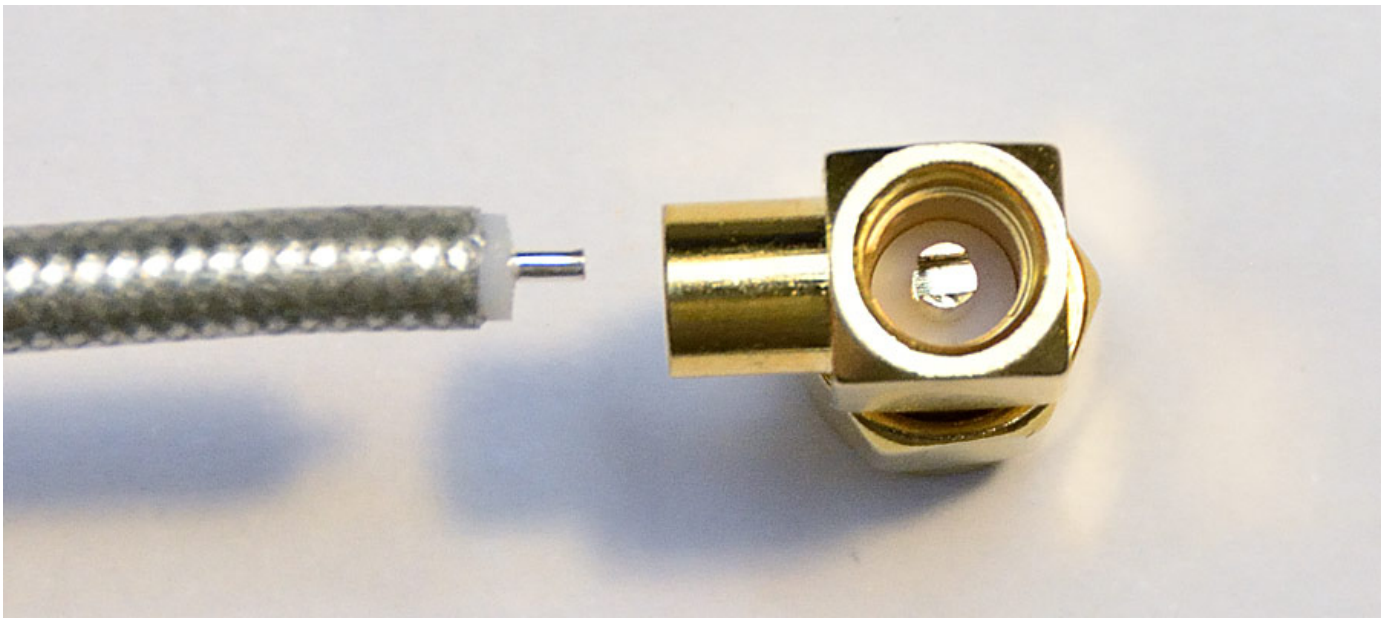
Working with 141 and SMAs is almost idiot proof.
141 has a teflon dielectric, so unlike RG58 type cable for CB or ham use, won't deform/melt with abuse from a soldering iron!

Start by preparing one end to suit soldering to the SMA (m) R/A connector.,

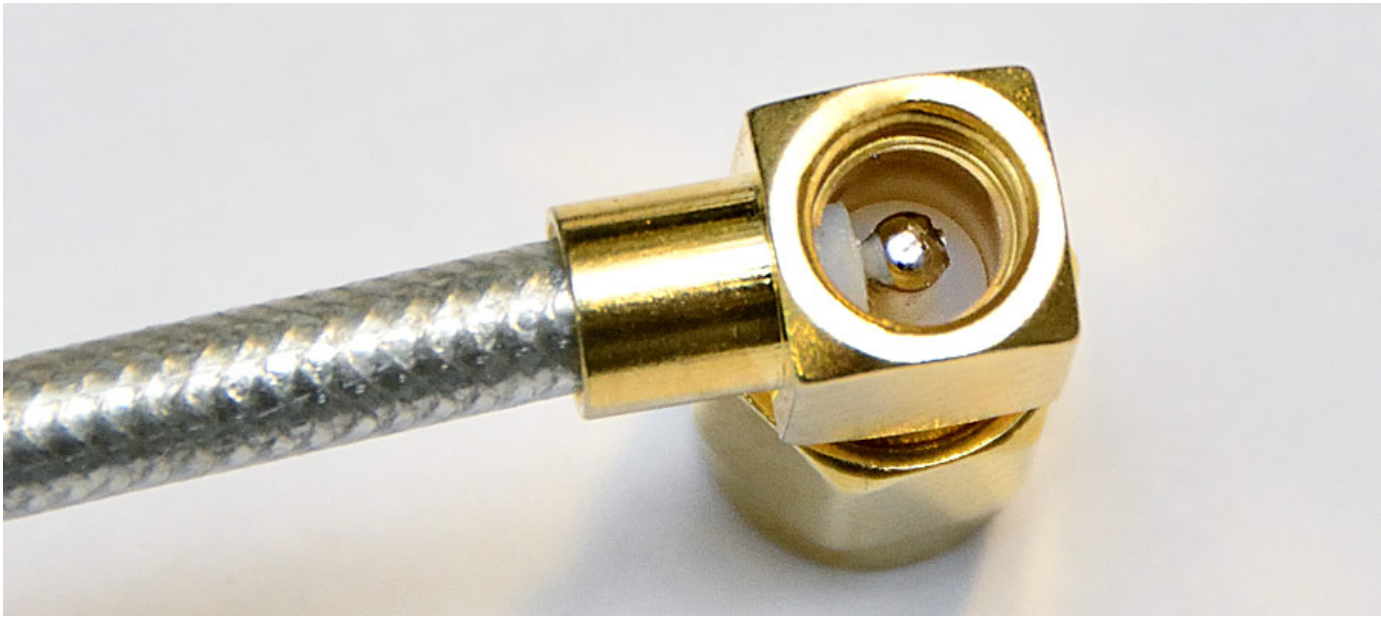
Score the outer sheath with a blade, as you rotate the coax (being careful not to cut into the dielectric):

Using pliers, twist and pull off the outer sheath, like so:
(I'm cheating, using a picture from further down)

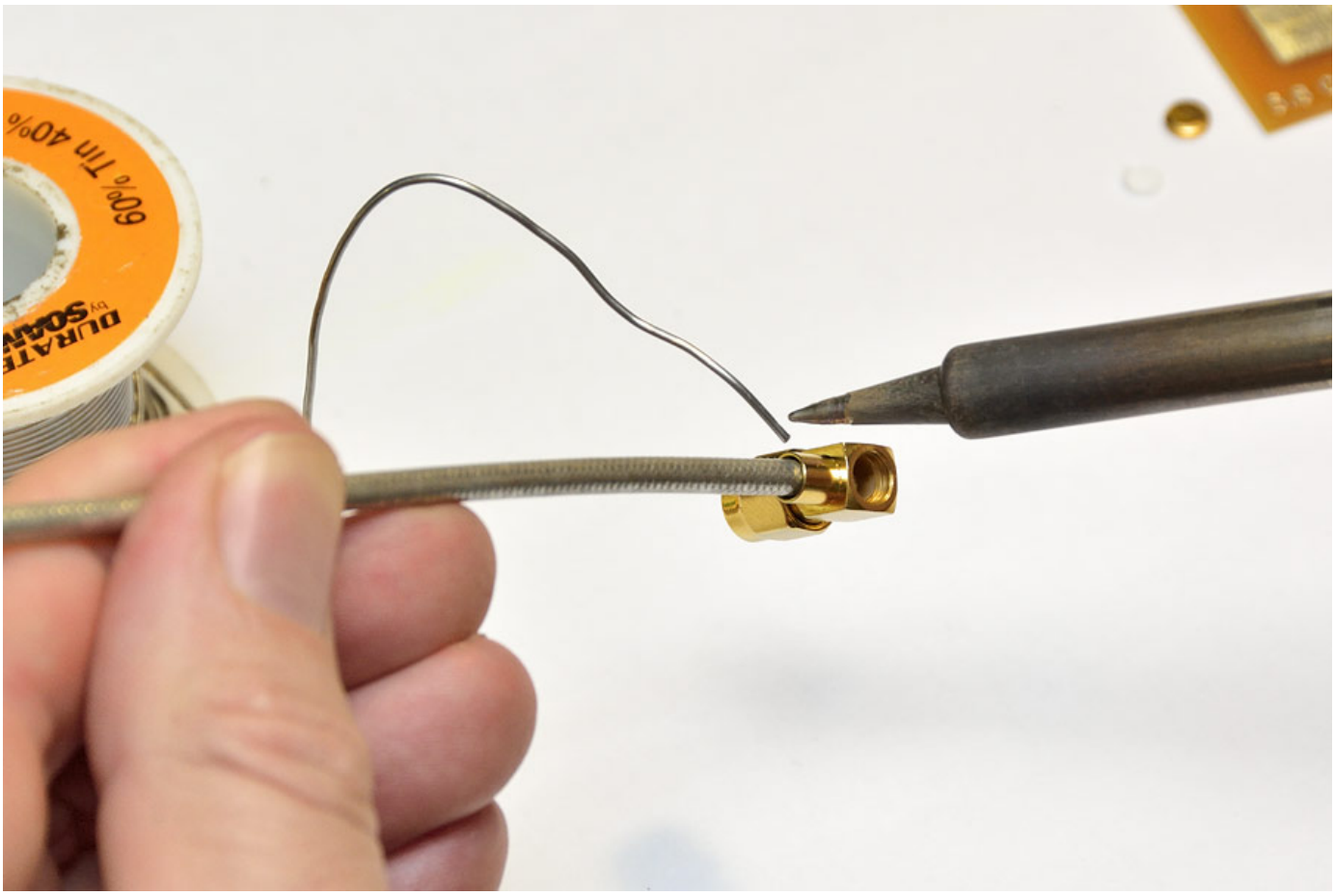
Coax end prepared, ready to solder to R/A SMA (m) connector:



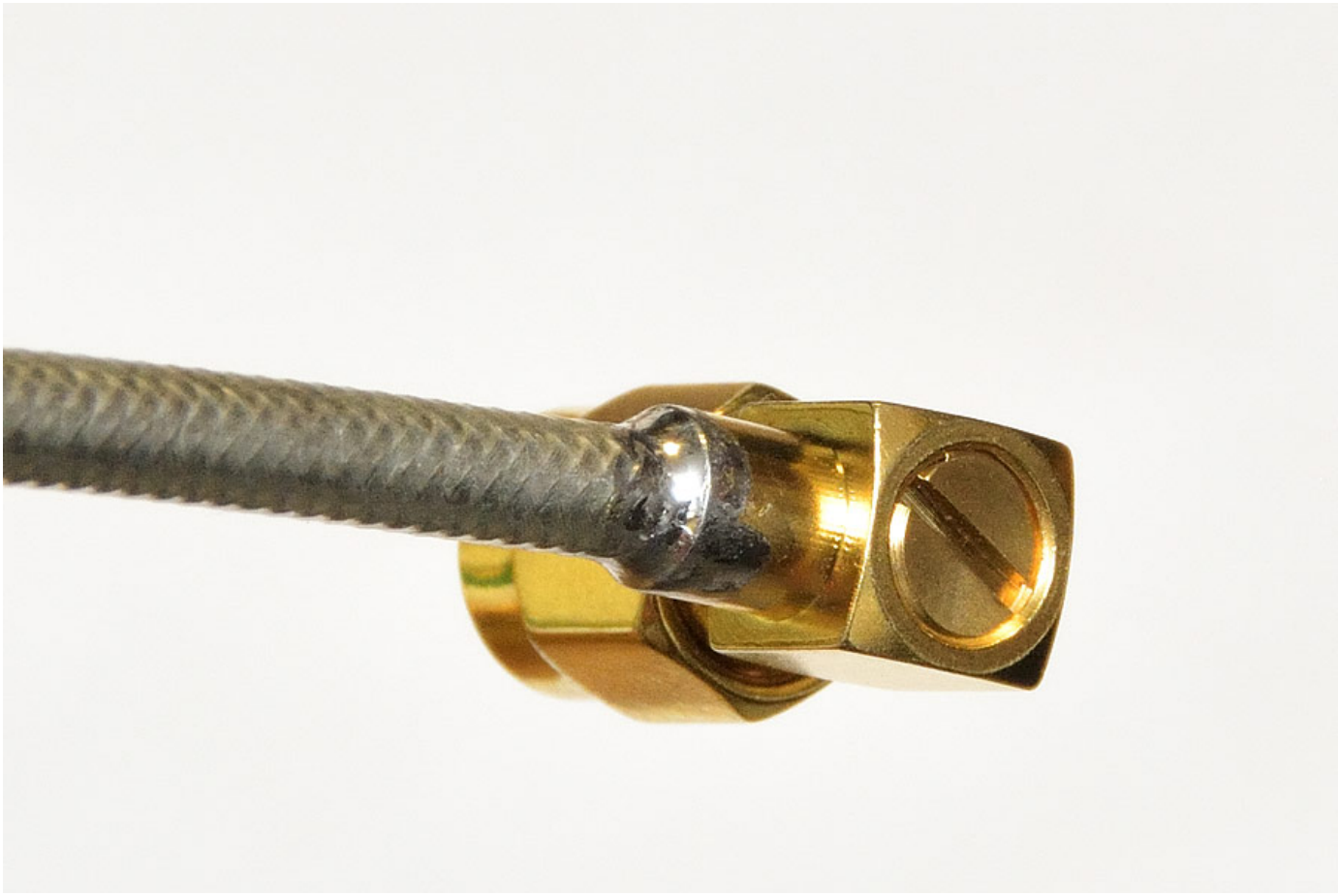
Coax inner soldered to centre pin:



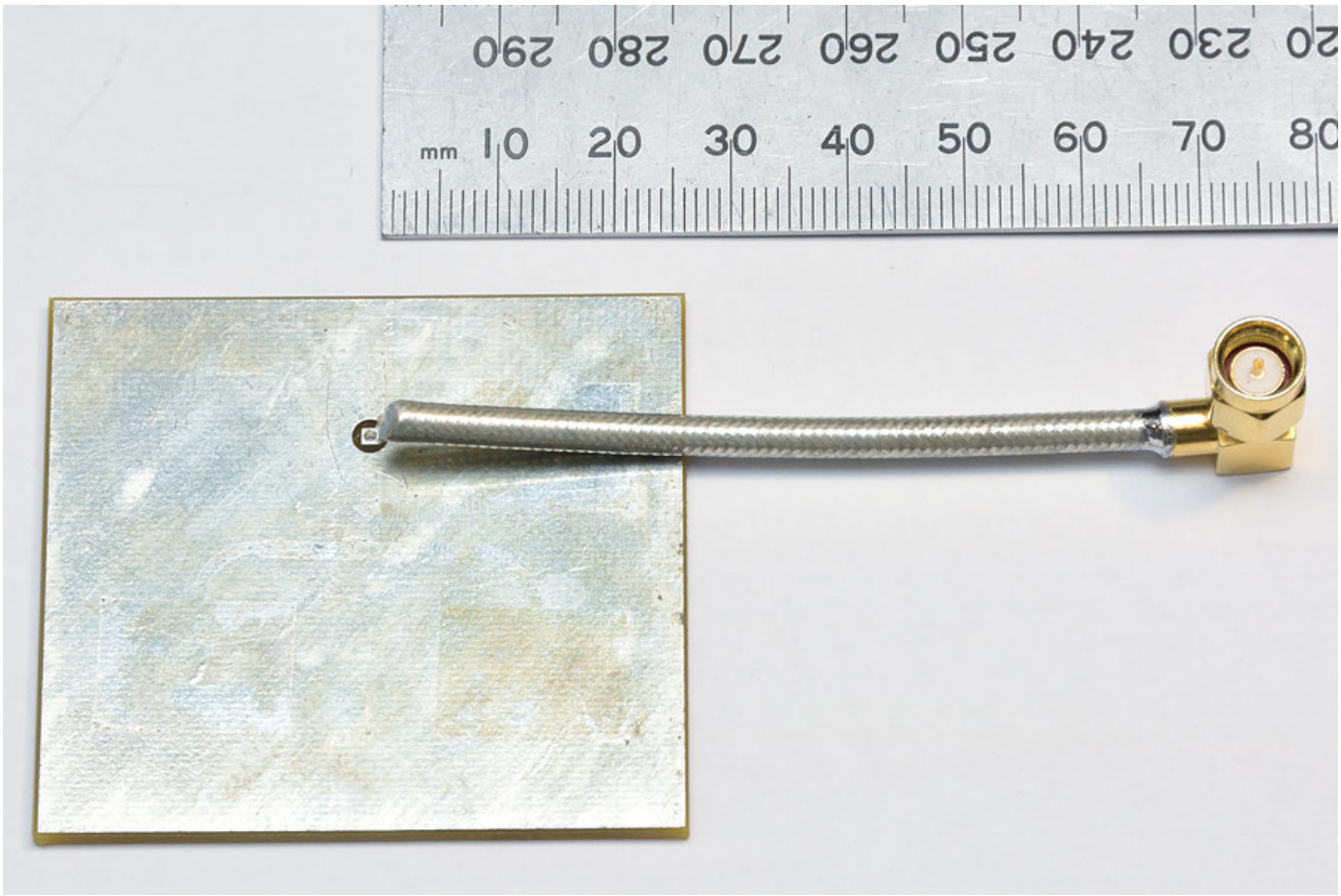
Add fillet of solder around outer sheath/connector body:



End cap screwed on:
Terminated!



A nominal length of coax is chosen:

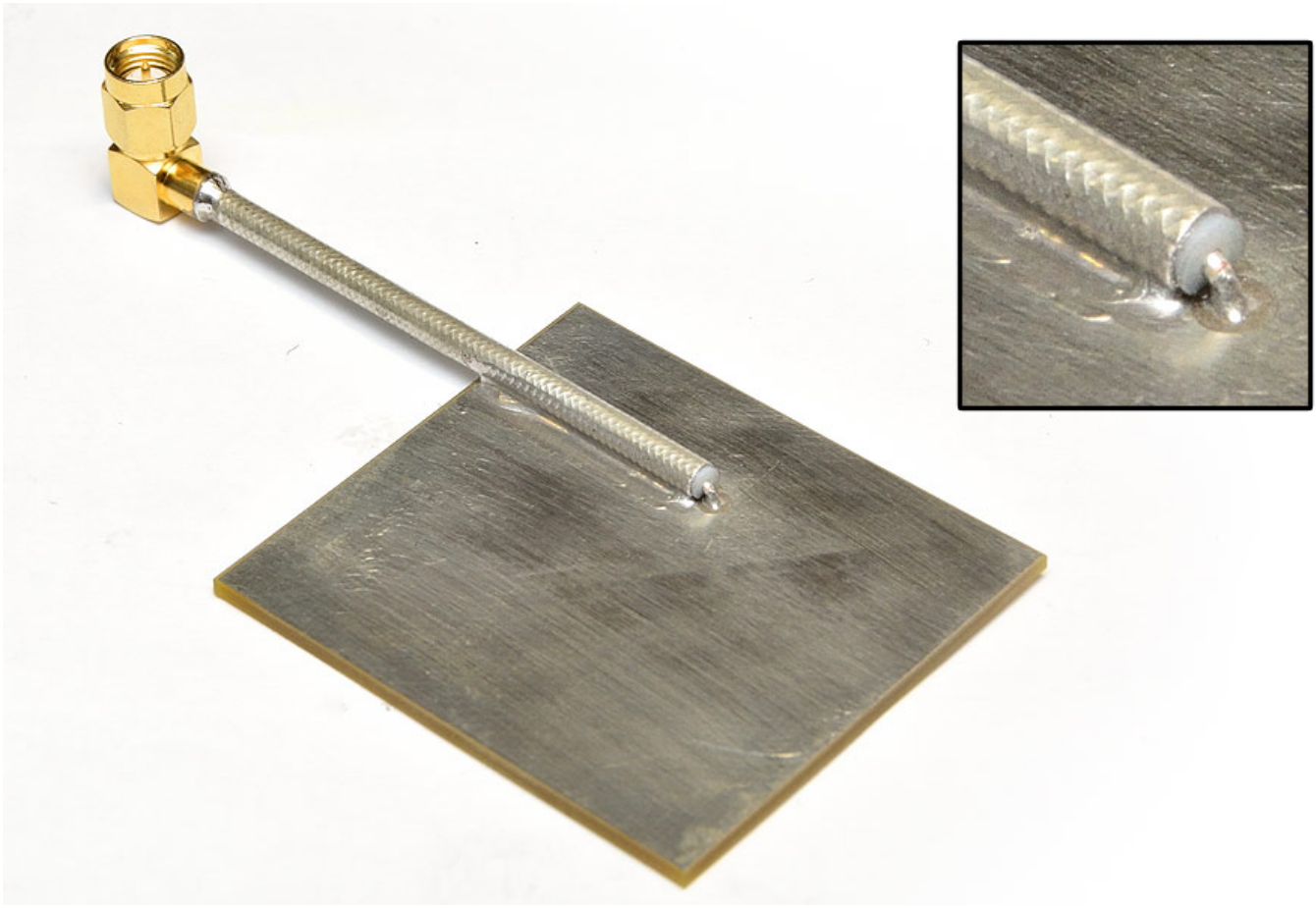


Sorry Americans, in millimetres… about time you adopted the metric system, like everywhere else! 😉

By the way, nothing fancy in the soldering department… just a bog standard Weller:

Coax end prepared, ready to solder to PCB:

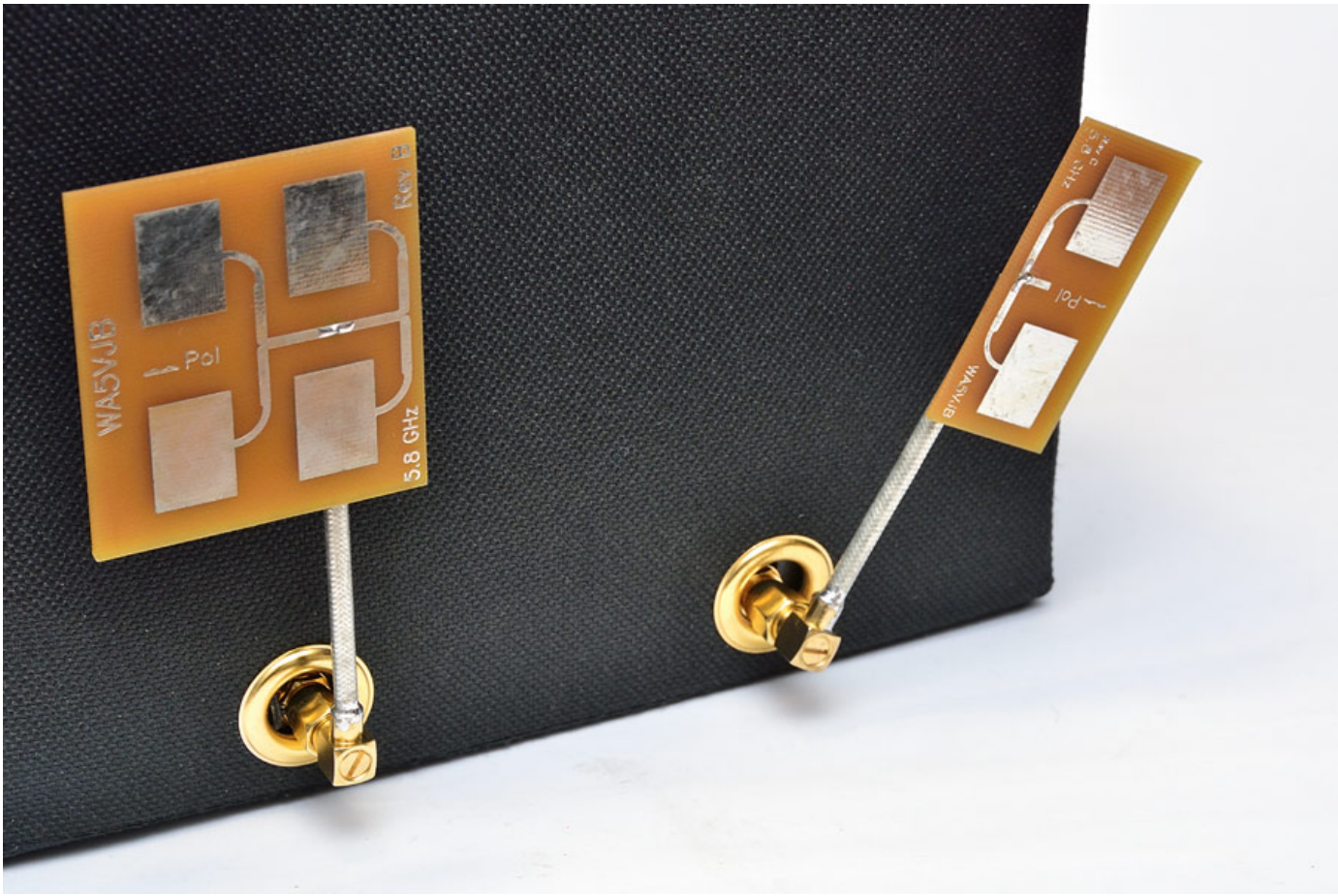
Coax soldered to the PCB antenna:

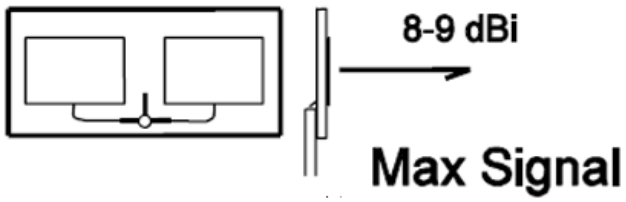


Note the fillet of solder that runs along the coax (on both sides), fastening the outer sheath to the ground plane of the PCB.

Installed on the rear of the 5.8 GHz diversity receiver/monitor, set up for horizontal polarisation:

A dual patch antenna was constructed as well.
Same procedure:





WA5VJB

Also on hand is a short in-line flexible “adapter”, with a male SMA at one end, a 30mm length of 141, and a female SMA, should a small angle be required:

As all hams operating on the microwave bands know, when dealing with SMAs, a 5/16″ spanner is always a handy tool to have:

Super Cheap Auto often have a bargain $1 bin on various size spanners.
Have a look for a 5/16″ next time you're in SCA.

« Hobby King T-Style CF Propellers review (http://vk4ghz.com/hobby-king-t-style-cf-propellers-review/)

5.8 GHz patch FPV antenna gain (http://vk4ghz.com/fpv-quad-patch-antenna-gain/) »

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Research Article

A Feed Network for a Novel E-H Shaped Microstrip Patch Antenna Array

N. Misran, M. Tariqul Islam and K.J. Ng

ABSTRACT

This research presents a beamforming feed network for a novel E-H shaped microstrip patch antenna array. The broadband microstrip antenna array was developed based on the novel L-probe fed inverted hybrid E-H (LIEH) shaped element design arranged into 4x1 uniform linear array antenna, fed by a feed network. The corporate 4-ways Wilkinson power divider which was fabricated in-house covering the frequency range from 1.83-2.29 GHz (460 MHZ) were integrated to form the beamforming feed network. The beamforming feed network comprises of commercial variable attenuators (KAT1D04SA002), variable phase shifters (KPH35OSC000) and the corporate 4-ways Wilkinson power divider which was developed in-house. The developed 4-ways Wilkinson power divider achieves bandwidth of 22.33% at VSWR 1.25. The measured isolation and insertion loss are 30.72 and 0.32 dB correspondingly. The feed network developed and fabricated in-house has been shown to function well within its design ability.

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INTRODUCTION

The evolution of **wireless communication**s has a deep impact on the field of antenna design. One essential component of the antenna array is its feeding technique. In an antenna array multiple radiating elements are assembled in an electrical and geometrical configuration. The antenna array provides beamforming capability to extract the desired signal and simultaneously filtering out the unwanted interference signal and environmental noise.

In designing array antenna, dipole, horn, waveguide slot or microstrip antenna can be used as radiating elements. However such radiating elements are unable to meet the requirements to be aesthetic and light weighted if implemented on the cell sites. To be aesthetic, the radiating element of the antenna array is desired to be small in size, low profile, conformal and compatible with integrated circuit and etc. Therefore, the radiating elements are preferred to be fabricated by microstrip technology that offers its non-electrical characteristic advantages (Pozar and Schaubert, 1995). Microstrip antennas are low profile and light weight. They can also be made conformable and well suited to be integrated with microwave integrated circuit (Lau *et al*., 2006; Zhang and Wang, 2006). In terms of fabrication, such system offers simplicity, which allows mass production and cost-effective manufacturing as well as high performance.

A combination of X- and C-band slots and patches were investigated and it was concluded that C-band patch/X-band slot concept had the greatest merit. In that design, the C-band feed network was on the same plane with the patches while the X-band feed was behind the ground plane. A combination of stacked perforated L-band patch overlaid over an array of C-band patches to achieve shared-aperture L/C-

Translate was reported in Shafai *et al*. (2000). Employing similar feeding principles, the C-band feed network behind the ground plane was aperture-coupled to the patches. The L-band feed network was on the same plane with the lower L-band perforated patch. This idea was further developed and applied to L- and X-band array in Pozar and Targonski (2001).

In their studies, Targonski and Pozar (1998) noted that it was impossible to accommodate two linear polarizations for two separate frequencies on a single feed layer. Maci and Gentili (1997), adding that single feed network for two frequencies was the most critical problem to solve for a dual-band array. Current dual-band shared-aperture technology suggests employing the higher- frequency feed network behind the ground plane (aperture-couple) and the lower-frequency feed network above the ground plane (coplanar feed). This may be because higher frequency feed produces higher spurious radiation, due to shorter wavelength and longer layout length. In a small array, feed layout above the ground plane may not produce excessive spurious radiation. However, in a larger array, feed line radiation may not be ignored.

In this research, the beamforming feed network was constructed using a four-way Wilkinson power divider of corporate structure type and a phase shifter-attenuator networks, connected in series to the ports to examine the scanning capability of the array. Due to the high cost and complexity of the design for planar and high resolution array, the design focuses on the development of a four-element (4x1) array antenna.

BEAMFORMING FEED NETWORK

A feed network is integrated with the antenna array. The radiating element of the antenna array is based on the LIEH shaped Microstrip Patch Antenna (MPA) that is arranged in a 4x1 linear array configuration which is termed as Broadband Microstrip Patch Antenna Array (BMPAA). The antenna array is constructed using two dielectric layer arrangement where a thick air-filled substrate was sandwiched between top-loaded dielectric substrate or superstrate with inverting radiating patch and an aluminium ground plane. The antenna array is used contemporary design techniques namely, the L-probe feeding, inverted patch and slotted patch techniques to meet the design requirement.

The difference in amplitude and phase of the array is fed to each element by feed network. This causes the radiating field of each element combines constructively in the intended direction and cancel out each other at other undesired directions in the far field (Balanis, 1997). Thus higher power gain and more flexibility in controlling the shape of the beamwidth and sidelobe levels can be achieved from this method. Moreover, the feed network allows the radiation pattern to be varied for beam scanning without any physical antenna motion.

Arrays can be fed in many ways, but the two general types of feed arrangement for microstrip patch arrays are series feed and corporate feed. Microstrip patch array in a series feed network is fed by a single transmission line, whereas patch array in a corporate feed network is fed by multiple transmission lines. Series feed network is much easier to design and fabricate compared to the corporate feed network. However this method will result in progressive phase delay between elements thus making unsuitable for phase scanning. Another problem with series feed arrangement is the high VSWR that is caused by the additive mismatches at the various junctions between the elements (Zurcher and Gardiol, 1995).

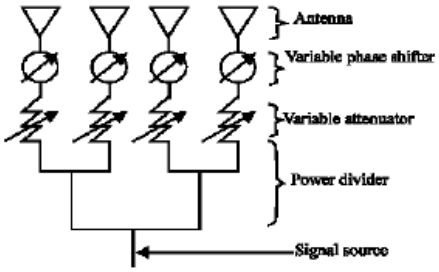

Fig. 1: Schematic diagram of a beamforming feed network

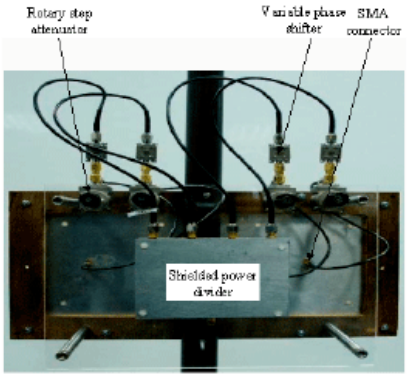

Fig. 2: Rear view of the 4x1 LIEH shaped BMPAA with a beamforming network

Corporate feed network is more flexible and offers better phase control over the performance of each array element. It is best suited for scanning phased array, multi beam arrays or shaped beam arrays. However, the use of corporate feed network is more preferable over

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s work as it is less affected by frequency scan. A corporate feed network supplies excitation individually to each array element and use equal line lengths and power dividers for each element.

The beamforming feed network for the BMPAA consists of a 4-ways corporate structure Wilkinson power divider and a network of phase shifter-attenuator as shown in Fig. 1.

The phase shifter-attenuator network comprises of a commercial off-the-shelf (COTS) variable phase shifter KPH350SC00 and step-rotary attenuator KATID04SA002, both are from KMW Inc. The 4-way Wilkinson power divider was developed and fabricated in-house by using Taconic RF-35 with the dielectric permittivity, ε_{r1} of 3.5, thickness h = 0.76 mm and with the dimension 25x20 mm. Table 1 gives the summary of the beamforming feed network design and its electrical parameters.

Figure 2 shows the fabricated beamforming feed network connected to the antenna array. As shown in the Fig. 2, the power divider is covered by an aluminium shield and the antenna array is behind the power divider.

Table 1: Summary of the beamforming feed network

Parameters	Values
A. Wilkinson Power Divider	
Substrate	Taconic RF -35 (ε_{r1} = 3.5, h = 0.76 mm)
B. Phase shifter	
Insertion loss	0.25 dB (1-2 GHz)
0.35dB (2-3GHz)	
VSWR	1.25
Phase shift	35° @ 2 GHz
C. Attenuator	
Frequency range	DC~3 GHz
Insertion loss	0.2 dB
VSWR	1.15
Attenuation step	10 attenuation step of 1 dB

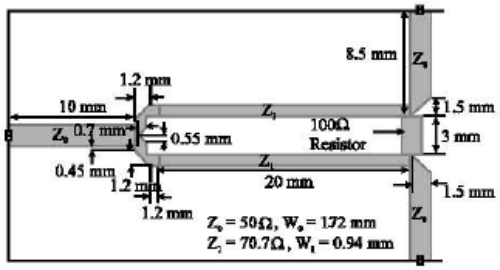

Fig. 3: The layout of 2-ways Wilkinson power divider on substrate Taconic RF 35 (ε_{r1} of 3.5, h = 0.76 mm and δ = 0.0018)

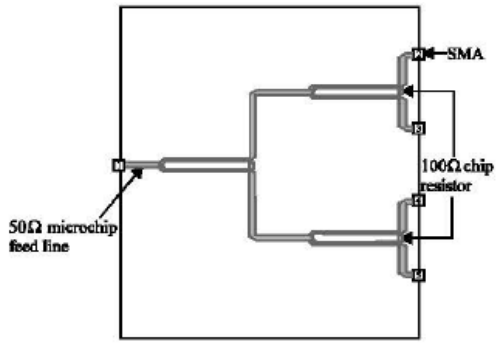

Fig. 4: The layout of 4-ways Wilkinson power divider

A 4-ways power divider is designed as a corporate feed network to feed the radiating elements with equal amplitude and phase. This power divider is based on Wilkinson power divider which offers good matching in all ports with good isolation between the output ports. The Wilkinson power divider has a useful property of being lossless when the output ports are matched, that is, only reflected power is dissipated (Pozar, 1990). The design schematic for the 2-ways Wilkinson power divider layout is shown in Fig. 3 and the layout of a 4-ways Wilkinson (3 dB) power divider is shown in Fig. 4.

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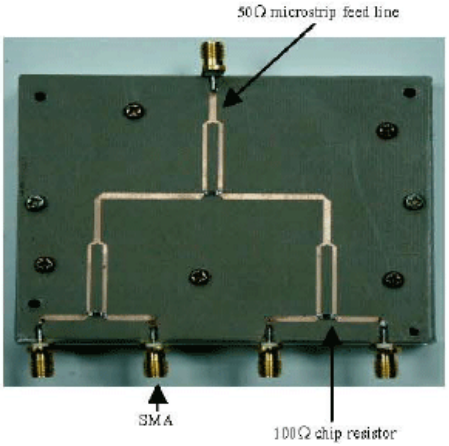

Fig. 5: Fabricated 4-ways Wilkinson power divider

Figure 5 shows the fabricated 4-ways Wilkinson power divider which is fabricated on Taconic RF-35.

RESULTS AND DISCUSSION

The antenna array and the power divider are simulated by *Sonnet® Suite em* simulator. The fabricated antennas and power divider are measured using the Agilent PNA E8358A network analyzer, Agilent ESG-DP Series E4436B signal generator, Advantest R3131A spectrum analyzer and the standard gain LPDA-0803 log periodic dipole antenna. SMA connector calibration kit, SMA male to male cable and automatic controlled rotator are also required for the measurement. Measurement is conducted in the UKM microwave lab and the UKM open field. The return loss measurement of 4-ways Wilkinson power divider is taken by using the Agilent PNA E8358A network analyzer. Return loss of every terminal is measured by terminating other element with 50 Ω load. Figure 6 shows the return loss at 1st terminal in 4-ways Wilkinson power divider. The measurement result shows higher value than the simulation results. This is due to the effect from SMA connector of every terminal. The aluminium ground plane is used to support SMA connector that is not been grounded properly. This installation problem causes the higher value of measurement; besides the used SMA connector, probe pin is not tapered. Simulated and measured return loss results are in good agreement in all terminals. The operating range for power divider is from 1.83-2.29 GHz (460 MHz) with bandwidth 22.33% at VSWR 1.25 considering all terminals.

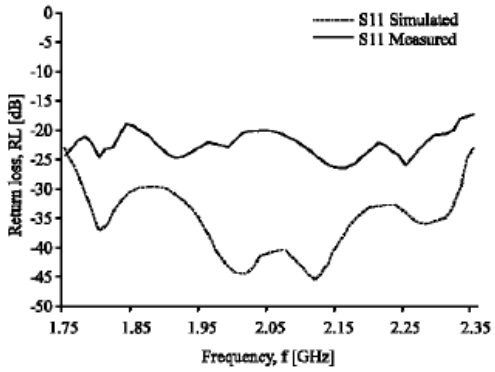

Fig. 6: Measured and simulated return loss of port 1 of the 4-ways Wilkinson power divider

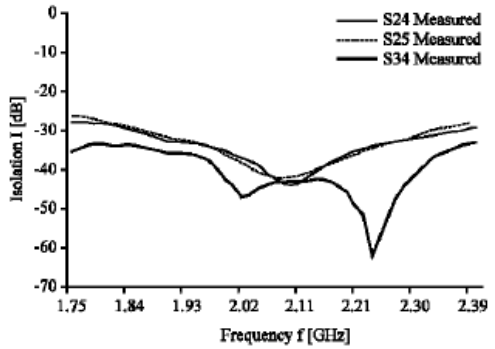

Fig. 7: Measured isolation of the 4-ways Wilkinson power divider

Table 2: Summarized result of power divider

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Properties	Measurement results
Bandwidth	460 MHz (1.83-2.29 GHz at VSWR = 1.25)
Isolation	30.72 dB
Insertion loss	0.32 dB
Amplitude balance	0.70 dB
Phase balance	5.40°

Isolation of 4-ways Wilkinson power divider is shown in Fig. 7. The maximum value of isolation in the operating frequency range is at 30.72 dB. The measurement results of power divider are shown in Table 2.

Figure 8 and 9 show the E-plane and H-plane radiation patterns of the antenna array at 1.91 and 2.14 GHz (resonance frequencies) by using feed network. Chebyshev amplitude distribution is used to have narrow beamwidth and low equal side lobe level (-20 dB). Chebyshev distributions value is shown in Table 3 as normalized in dB.

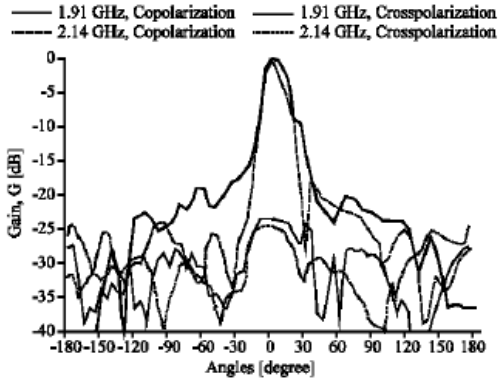

Fig. 8: Measured E-plane radiation patterns for the LIEH shaped BMPAA at 1.91 and 2.14 GHz

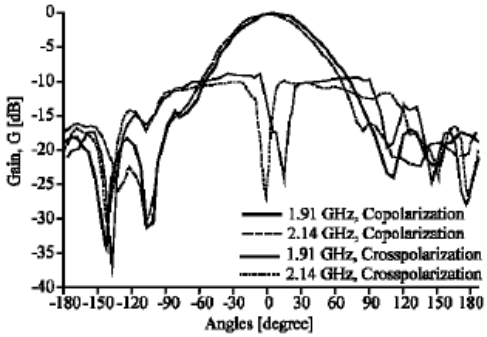

Fig. 9: Measured H-plane radiation patterns for the LIEH shaped BMPAA at 1.91 and 2.14 GHz

Table 3: Chebyshev distributions for each element of 4x1 BMPAA

Element-1	Element-2	Element-3	Element-4
0.48312 (3.1594 dB)	1(0 dB)	1(0 dB)	0.48312 (3.1594 dB)

Table 4: Amplitude and phase value for measurements

0° Scanning Element-1	Element-2	Element-3	Element-4
-10.06 dB	-6.5 dB	-7.5 dB	-9.65 dB
100.0°	-50.85°	149.3°	2.55°

However, the exact amplitude and phase distribution were unable to obtain due to the limitation of the available variable attenuator and phase shifter. Therefore, the antenna radiation pattern result is expected to be distorted from the desired performance.

Scanning pattern is measured for 0°.Amplitude and phase values are chosen by considering the following initial phase angle and by employing progressive phase shift equations. Amplitude and phase values for measurement are shown in Table 4.

Initial phase angle: Port 1: 100°
Port 2: 105°
Port 3: 101°
Port 4: 105°

The phase in each element of the array is progressively shifted by $\alpha = -\beta d\cos\theta$, where β is the phase shift factor and d is the inter-element spacing.

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patterns show some fluctuations due to the reflection from some obstacles in the field, however, they have good beam patterns and crosspolarization level, which is shown in Fig. 8 and 9. The 3 dB beamwidth is closed to 25° for the E-plane while the 3 dB beamwidth is about 65° for the H-plane. The H-plane radiation pattern is virtually symmetrical, while the E-plane radiation pattern exhibits some asymmetries, which is similar in the report by Huynh and Lee (1995) using a thick substrate. As shown in Fig. 8, the sidelobe levels are unequally distributed. The first side lobe levels at 1.91 and 2.14 GHz are -18.9 dB (at -60°) and -20.47 dB (at 45°), respectively. This result is due to the amplitude/phase unbalances in the beamforming feed network. The maximum crosspolarization of the array is in the order of -23.47 and -10 dB for E-plane and H-plane, respectively.

CONCLUSION

The beamforming feed network was constructed using a four-way Wilkinson power divider of corporate structure type and off the shelf variable phase shifter (KPH350SC00@ KMW Inc) attenuator (KATID04SA002@ KMW Inc) networks to measure the beamforming capability of the array antenna. Due to the high cost and complexity of the design for planar and high resolution array, the design focused on the development of a uniform four-element (4x1) array antenna. The array designed by employing novel hybrid E-H shaped design, inverted patch, slotted patch and L-probe. The developed 4-ways Wilkinson power divider enjoys the bandwidth of 22.33% at VSWR 1.25. The measured isolation and insertion loss are 30.72 and 0.32 dB correspondingly.

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Mobile Phone Antenna Design

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Abstract

This thesis focuses on mobile phones antenna design with brief description about the historical development, basic parameters and the types of antennas which are used in mobile phones. Mobile phones antenna design section consists of two proposed PIFA antennas. The first design concerns a single band antenna with resonant frequency at GPS frequency (1.575GHz). The first model is designed with main consideration that is to have the lower possible PIFA single band dimensions with reasonable return loss (S11) and the *efficiencies. Second design concerns in a wideband PIFA antenna which cover the range from 1800MHz to 2600MHz. This range covers certain important bands: GSM (1800MHz & 1900MHz), UMTS (2100MHz), Bluetooth & Wi-Fi (2.4GHz) and LTE system (2.3GHz, 2.5GHz, and 2.6GHz). The wideband PIFA design is achieved by using slotted ground plane technique. The simulations for both models are performed in COMSOL Multiphysics.*

The last two parts of the thesis present the problems of mobile phones antenna. Starting with Specific absorption rate (SAR) problem, efficiency of Mobile phones antenna, and hand-held environment.

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The work of the thesis has been carried out of Blekinge Institute of Technology (BTH), Karlskrona, Sweden under the supervision Dr. Benny Lövström.

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Acronyms

LAN	Local area network.
GSM	Global system for mobile communication.
PIFA	Planar inverted – F Antenna.
UMTS	Universal mobile telecommunication system.
DECT	Digital enhanced cordless tele communication.
PDC	Personal digital cellular.
RFID	Radio frequency identification.
HPBW	Half – power beam width.
FNBW	First null beam width.
IFS	Iterative function system.
SAR	Specific absorption rate.
RF	Radio frequency.
PMA	planar monopole antenna.
MOM	Method of moments.
FDTD	Finite difference time – domain.
ESA	Electrically small antenna.
Wi-Fi	Wireless fidelity
GPS	Global positioning system
RHCP	Right Hand circularly polarized
IFA	Inverted F-antenna
PTFE	Polytetrafluoroethylene
PEC	Perfect electrically conductor
LTE	Long term evaluation

Chapter1

1.1 Introduction

The huge development of the mobile phones have grown up rapidly in the last years, frequency bands have come up and the market is asking for smaller mobile phones with more services which give the user the ability to use the mobile phone with good signal performance and helps to use the mobile phone around the all world. On top of that it is important to reduce the risks affecting in the human body because of the antenna radiation.

In the past the mobile phone was so heavy, big and had external antenna on the top of the phone which effect badly on the human head and most of the signal radiate is reflected and absorbed by the human head which lead to bad efficiency.

Nowadays the internal antenna has been using instead of the external antenna the main reason of that is the internal antenna has a good relation with SAR rate, on the other hand the size of the phone became smaller.

Recently there are many types of the internal antennas for example PIFA antenna (Planar Inverted- F Antenna), fractal antenna and monopole antenna. Those kinds of antennas can cover a single band, dual band, wideband and multiband based on the design of the antenna.

PIFA antennas are used widely in mobile phone antennas design due to its advantages such as SAR rate and less interaction with hand-held environment, but one of the significant problems of PIFA antennas that's PIFA antennas have a narrow bandwidth. Wideband and multiband PIFA antenna can be a solution of that problem. By designing multiband antenna or wideband antenna

it's possible to have one antenna can cover more frequency such as using one antenna that can cover the very important bands which are in use in most countries GSM, UMTS, Wi-Fi and LTE.

1.2 Thesis Motivation

The work in this thesis is motivated by the necessity to investigate antenna structures which can be integrated in today is mobile phones which have a small available space for antennas inside. Nowadays the demand on low profile antennas increases rapidly that's in response to the huge developments of mobile phone devices either in functions or in size.

The other motivation is to investigate and design a wideband PIFA antenna. In general PIFA antennas have a disadvantage that's PIFA antennas have a narrow bandwidth, but on other hand PIFA antennas have many advantages such as best SAR rate, low profile and easy to be integrated inside mobile phones which make PIFA antennas the best candidate for mobile phones antennas. There are many papers published which study and discuss the narrow bandwidth of PIFA antennas, but most of those papers give the solution of the problem in dual band PIFA or in multiband design which in both cases do not solve the problem totally. Since there are some bands remain without covering, consequently the need of several antennas appears for covering low and high frequency bands.
So this thesis presents a study of a proposed wideband PIFA antenna with bandwidth more than 30% and the bandwidth cover an important range which could cover GSM, UMTS, and some bands of LTE system.

1.3 The Development of Mobile Phone Antennas

In the first generation 1G of mobile system, the system was operating at 800 MHz and as known it was analog. The first antenna handset for one quarter of wavelength has length about 9.4 cm, and it was one antenna only. The first mobile phone was Motorola DynaTAC8000X[1] with type

antenna is sleeve dipole as shown in the figure 1.3.1 which is not use any more in modern design of mobile phones antennas, however it still is used in different wireless LAN access points.

Sleeve dipole antenna has an efficient performance; the length of it is about half the wave length at its frequency. So at 850 MHz the antenna should be 176mm.

With the improvement on mobile phone and due to the dramatic minimizing size of the mobile phone, there were no needs for a sleeve dipole to be proportional to the cellular phone.

Photo was loaded from http://mashable.com/2014/03/13/first-cellphone-on-sale/

Figure 1.3.1 Motorola DynaTAC8000X with Sleeve dipole antenna.

In 1990s 2G launched which offered new services such as text message, and it is operated at GSM 900 MHz, where later on at 1800 MHz unlikely 1G, the second generation handset antenna has two antennas monopole and helix with only single band supporting as shown below in figure 1.3.2 and 1.3.3.

http://tech-kid.com/nokia-phone.html

http://www.northstandchat.com/showthread.php?289468-Your-first-ever- Mobile -Phone-and-what-make-and-model-was-it-!/page3

Figure 1.3.2 Nokia 1011 supports only GSM900 single band band.

Figure1.3.3 Motorola m300 support only GSM1800 single

In 1997, Motorola produced a mobile device Motorola mr601 which was the first dual band GSM phone, it supported GSM900 and GSM1800 dual band and its antenna consists of two antennas helix antenna which has travelling wave in the shape of corkscrew with circularly polarized and whip antenna which can be consider as dipole antenna and it is Omni-directional radiation pattern. That's model phone offered the ability to access network in over 70 countries by the end 1997.

In the first dual band PIFA operating at GSM900 and GSM1800, invented by Prof. Peter Hall in 1996 U.K, the first dual band PIFA a slot with a certain geometric dimensions, lead to support two different bands.

The development timeline of mobile phone has several significant steps, one of them was in 1999, when Nokia launched Nokia3210 as shown in figure 1.3.4 the first mobile with fully internal antenna, support both GSM900 and GSM1800 dual band, and was one of the most popular handset with over 160 million being sold [2].

In 1999, Prof. Peter Hall comes back with the first Triple Band PIFA which operates at GSM800/1800/1900. He designed Triple Band PIFA with two slots in the ground plane of the PIFA antenna, that's let the antenna to transmit and receive different band of frequencies.

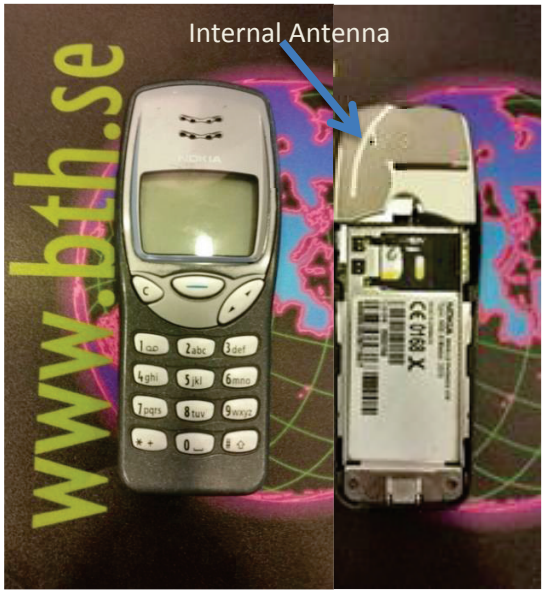


Figure 1.3.4 Nokia 3210 and its internal antenna which is installed to the top left of the back view of the mobile phone.

The first commercial 3G network was launched by Hutchison Telecommunications branded as 3G in June 2003, and in 2004 Nokia released Nokia6630,as showing in figure1.5 which was the first mobile phones allows truly global roaming, and first mobile with dual mode tri-band, GSM900/1800/1900 and UMTS 2100.

The photo to the left was loaded from http://www.mondomobileblog.com/2010/03/31/codici-segreti-nokia-6630/

Figure 1.3.5 Nokia6630 mobile and its internal antennas.

Table1.1 The timeline of development antenna use in mobile phones [1].

Year	1900	1950	1970	1990	2000	2010
Frequency	< 10 kHz	< 30 MHz	< 800 MHz	< 1.9 GHz	< 2 GHz	3 GHz 5 GHz
System	Telegraph / telephone for train, ship, police cars	Vehicle and portable systems for business, Pager	Mobile phones (analog), cordless phones, pagers, GPS	Mobile phones (digital),aircraft (voice and data) personal phone	Mobile phones (multimedia), Wireless access, Bluetooth	Mobile phones (high data rate), Ultra wideband,
Antenna	Monopole dipole, whip,	Blades, ferrite coil, helical antenna	Corner reflector, PIFA, helix	Meander line, normal mode helix, ceramic chip, adaptive array	Wideband, multiband built-in antennas, adaptive array, MIMO	Small compact functional antennas, implant antennas

Table 1.3.1 shows the time line of development in both sides' frequencies and antenna which are used for mobile phones systems. The frequencies are raised from 10 kHz till 3 or 6 GHz which is considered as a 4G system.

Cordless phones navigation systems and multichannel access systems have contributed dramatically to the development of antennas design, while the personalization of mobile terminals raised the needs of small antennas. Other needs for small antennas are hybrid systems, the handset of multiband GSM combined with DECT (Digital Enhanced Cordless Telecommunications) for example.

The specifications of the handset small antennas are low profile, compact and lightweight, which has led to modify and develop the antennas. One of the antennas is a modified PIFA, which has changed that it is difficult to recognize it as an original PIFA. The modified PIFA was first used in GSM handsets and has become more common in several different mobile terminals. Other modified antennas such as monopole were used in PDC (Personal Digital Cellular) with a normal mode helical antenna.

In modern mobile phone, we can conclude a summary of the considerations in design process as following:

- The requirement of multiband and built–in antenna.
- The requirement of small size antenna due to small size mobile terminals, for example RFID units (Radio-frequency identification).
- The requirement of function antennas such as adaptive antennas and MIMO for transmission high data rate.

Printed antenna has been used in mobile terminals; also Printed antennas have the smallest dimensions comparing with other installed small antennas. Chip antennas or a printed antenna consists of dielectric material and a ground plane which is a part of the antenna. Chip antennas offer a high gain, appropriate radiation pattern, and high efficiency.

Chapter 2

Basic Mobile Phone Antennas Principles

2.1 Parameters of Mobile Phone Antennas

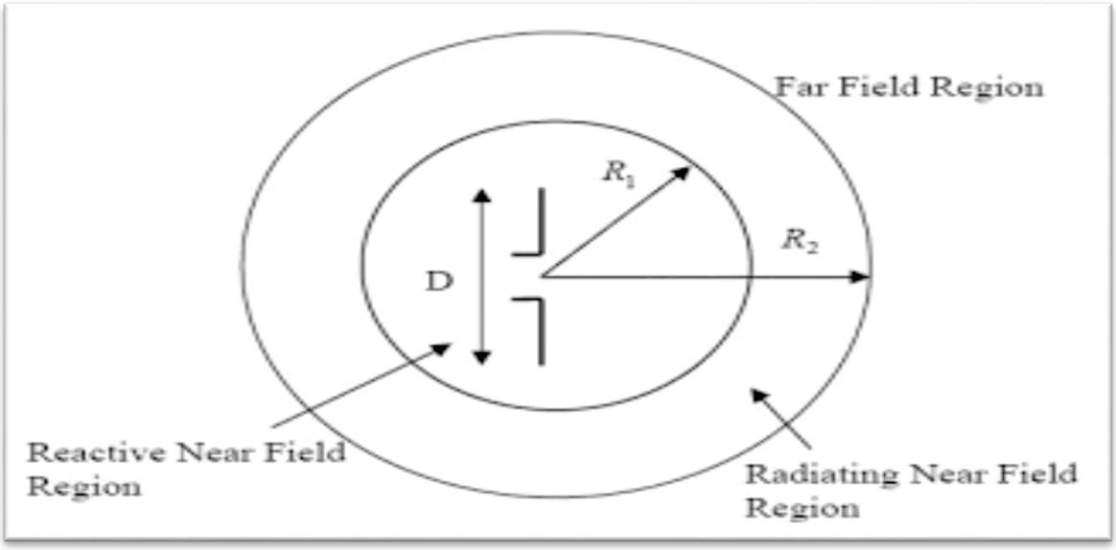
Radiation pattern

Radiation pattern can be describe by power radiated through the antenna, which depends on the direction, and it can be represented as graphical or mathematical function of the radiation properties.

There are two common types of antenna pattern , the first one is a power pattern and the second is field pattern , the first type is considering the power radiated and the position of the spherical coordinate and the second type which depend on the field if we have electrical field or magnetic field with the position of spherical coordinate.

The field region of the antenna pattern can be dived to three regions. The first region is Reactive Near Field Region which is the closest region to the antenna so it gives possibility to control the

energy and standing waves. The second region which is locate between the Reactive near field region and the far field region where control the radiation field, The last region is the far field region which has a far distance of the antenna , as showing in figure 2.1.1.



http://antenna-practices.blogspot.se/2008/11/antenna.html

Figure 2.1.1 The radiation pattern and the field regions.

There are many types of antenna pattern

1- Isotropic pattern the radiation of the isotropic is symmetric in all directions which is the ideal antenna but it is not found until these days.
2- Directional pattern which radiate more focusing with one direction.
3- Plane pattern which consist of two principles the electrical field plane (E-plane) and the magnetic field (H-plane).

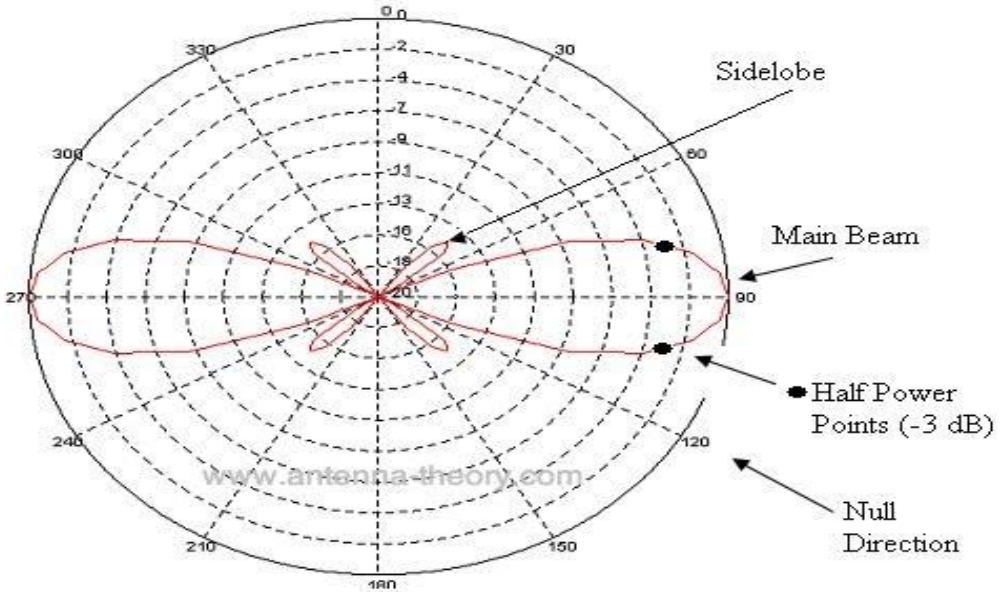
What is (E-plane) and (H-plane)?

E-plane consists of electrical field vector combined with the direction of maximum radiation.

H-plane consists of magnetic field vector combined with the direction of maximum radiation.

Parameters of antenna pattern

1- Main lobe is the radiation lobe going in the same direction of the maximum radiation.
2- Side Lobe is the radiation lobe of any directional lobe except the intended direction.
3- Back lope which has the opposite direction of the main lobe.
4- Half-Power Beam width (HPBW) the angular between two directions the half power point and the beam width which equal to -3db.
5- First Null Beam width (FNBW) - angular width between two directions the first zero and other side in the side lobe.



http://www.antenna-theory.com/basics/whyantennasradiate.php

Figure 2.1.2 A polar figure of an antenna with performance Parameters.

Directivity

Directivity is one of the antenna parameters that guide us to measure the direction of the radiation pattern, the directivity help us to determine the efficiency of the antenna, directivity of isotropic antenna equal to 1 (0 dB).

The directivity can be calculated by determining the maximum value of the magnitude of the radiation pattern which is defined by F(θ, ϕ) and the average power of all the direction which is equal to 1 so, the formula of calculating the directivity it can be written as:

$$\boldsymbol{D = \frac{1}{\frac{1}{4\pi}\int_0^{2\pi}\int_0^{\pi}|F(\theta,\phi)|^2 sin\theta d\theta d\phi}} \quad \textbf{(2.1.1)}$$

θ = The angle between the z-axis and the vector from the origin to the point

ϕ = The angle between the x-axis and the projection onto the x-y plane.

Mobile phone antenna should have a low directivity, because is not useful to use an antenna with high directivity, that is the received signal it can be from different directions. On the other hand the base station needs a high directivity.

Table 2.1.1 Directivity of different types of antennas [4].

Antenna Type	Typical Directivity	Typical Directivity (dB)
Short Dipole Antenna	1.5	1.76
Half-Wave Dipole Antenna	1.64	2.15
Patch (Microstrip) Antenna	3.2-6.3	5-8
Horn Antenna	10-100	10-20
Dish Antenna	10-10,000	10-40

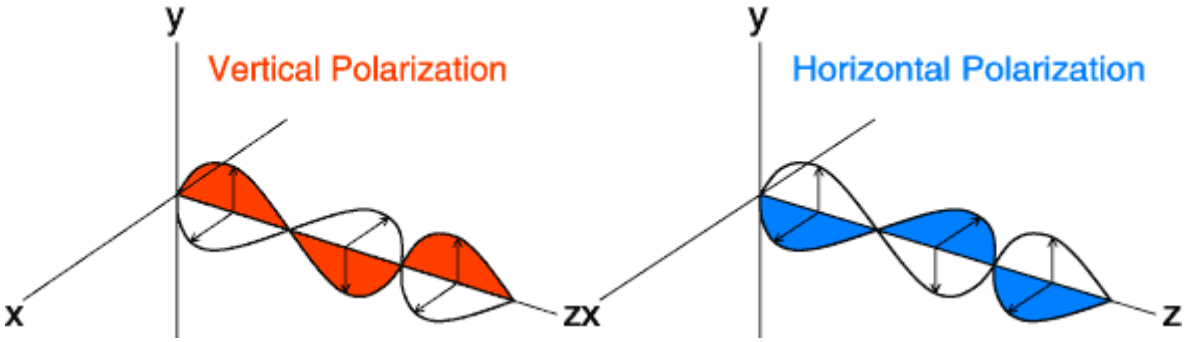
As showing in the table 2.1.1 above the directivity of different types of antenna so, the properties of the antenna have to be chosen regarding to what we need, as we have mentioned if you want to choose antenna for mobile station as cell phone, car or computer we will choose antenna with low directivity.

Polarization

One of the characteristics for any types of antennas is the direction of the electrical field vector of the electromagnetic wave. The polarizations of electromagnetic wave depend on the electrical field, which is perpendicular with the magnetic field.

The polarization of antennas can be classified in three different types based on electrical field

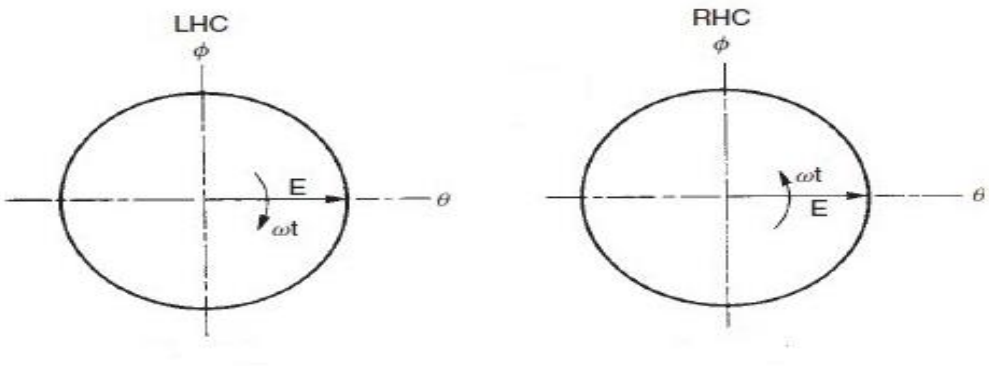
1- **Linear polarization:** as seen in the figure 2.1.3 the electrical field travels with the same direction of the plane wave. There are two types of linear polarization:
   - **Vertical polarization:** the electrical field travelling orthogonally to the y-axis and the magnetic field is travelling horizontally to x-axis.
   - **Horizontal polarization:** the electrical field is travelling horizontally to the x-axis and the magnetic field is travelling orthogonally to the y-axis.



http://www.srh.noaa.gov/hun/?n=dualpol

Figure 2.1.3 Vertical polarization and horizontal polarization.

2- **Circular polarization:** the electrical field travels circularly as seen in figure 2.1.4. It can be divided to two types regarding to direction of the circle:
   - Right Circular polarization: the direction of the circle is clockwise.
   - Left circular polarization: the direction of the circle is anti-clockwise.

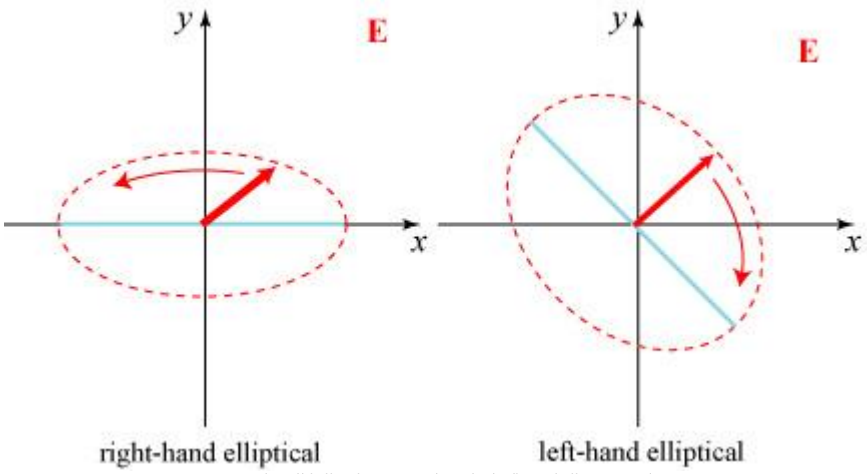


http://www.antetec.com/news/2010/01/27-371.html

Figure 2.1.4 Right circular polarization and left circular polarization.

3- **Elliptical polarization:** the electrical field travels with an elliptical shape based on the direction of the electrical field figure 2.1.5 it can be divided to two types:

- Right elliptical polarization which is has direction with clockwise.
- Left elliptical polarization which has direction anti-clockwise.



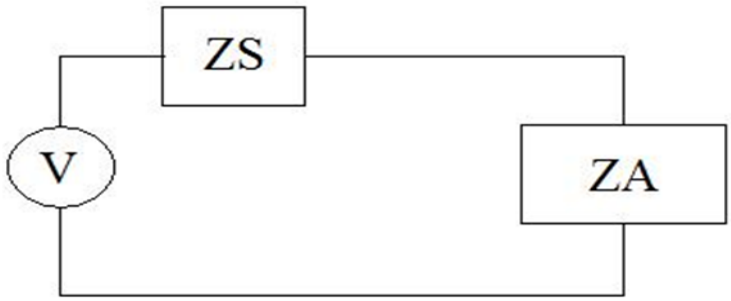
http://skullsinthestars.com/2012/09/23/hairy-balls-in-optics/

Figure 2.1.5 Right elliptical polarization and left elliptical polarization.

Impedance

The impedance of the antenna mainly depends on the input voltage and current for the power losses equal to the power transmission of the antenna. The benefits of knowing the input impedance it is lead you that to choose the suitable transmission line.

The impedance of an antenna ZA must be matched with its impedance transmission line ZS (feeder). Otherwise there will be amount of a reflection wave back to the signal source. To have maximum power at the antenna from source as shown in figure 2.1.6 ZS should be equal to ZA



http://www.antenna-theory.com/basics/whyantennasradiate.php

Figure 2.1.6 The equivalent circuit of an antenna and the transmission line.

Antenna gain

One of the antenna properties to describe the value of the power transmission to the peak radiation direction. The efficiency of the antenna always relate with the antenna gain so if the efficiency is perfect 100% that's mean the gain and the directivity are equal to each other, it can be describe by the following formula:

$$G = \varepsilon_R D \qquad (2.1.2)$$

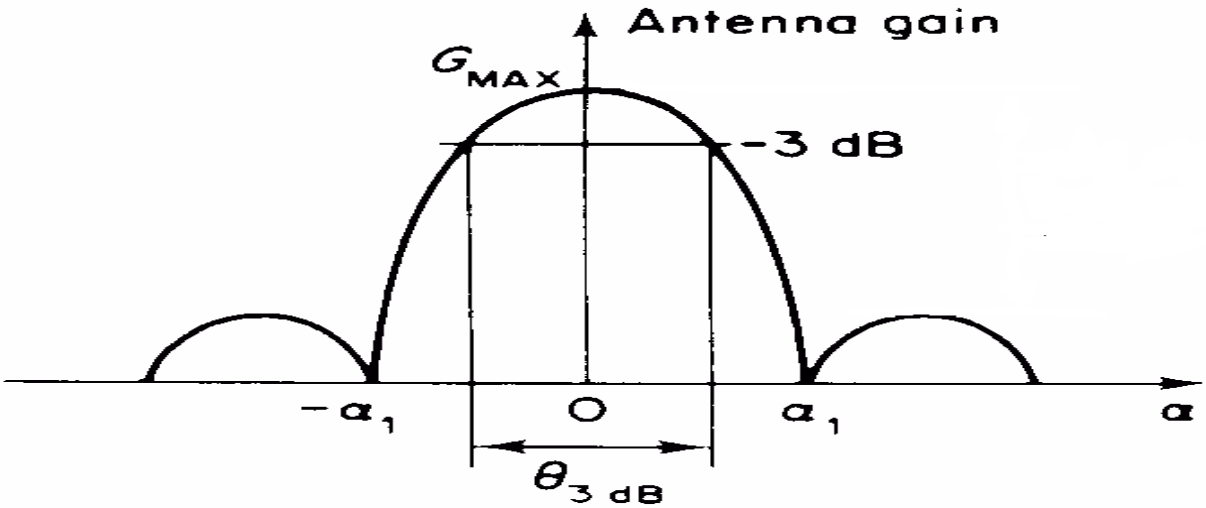
Antenna efficiency

The relation between the input power to the antenna and the dissipated power which is the losses power within antenna, regarding to the relation between the power radiate and the power input the efficiency of the antenna can be written by the following formula:

$$\varepsilon_R = \frac{P_r}{P_{in}} \qquad \textbf{(2.1.3)}$$

Antenna Beamwidth (θ)

The angle between the two points of the half power of the radiation pattern, the beamwidh could be clarify as the half power, the directivity become high if the beamwitdh very small. The beamwidth is found by determining the two points with 3db as showing in figure 2.1.7.



http://happy.emu.id.au/lab/rep/rep/9510/txtspace/9510_010.htm

Figure 2.1.7 Half power Beamwidth.

Effective area

To calculate the transmission power (P_t) and determine how much power has been lost due to the plane wave, regarding to the relation between the power density (P) and the effective area (A_e) the following formula can be written as:

$$P_t = PA_e \qquad (2.1.4)$$

To calculate the effective area we should consider the relation between the effective area and the antenna gain the formula can be written as:

$$A_e = \frac{\lambda^2}{4\pi} G \qquad (2.1.5)$$

Bandwidth

One of the other important parameters is bandwidth which determines the range of the frequencies where the antenna is operating, table 2.1.2 shows different common types of antennas with frequency range of each antenna. The percentage of the bandwidth can be calculated as following:

$$Bandwidth\% = \frac{f_{max} - f_{min}}{f_{Center}} \times 100 \qquad (2.1.6)$$

Table 2.1.2 different common types of antennas with frequency range of each antenna [4].

Antenna	Center frequency	Frequency range	Percentage bandwidth
Patch	1000MHz	985-1015MHz	3%
Dipole	1000MHz	960-1040MHz	8%
Horn	1000MHz	154-1848MHz	169.40%
Spiral	1000MHz	95-1900MHz	180.50%

VSWR

Voltage standing wave ration or S_{11} which describe the reflection power coefficient or the return losses of the antenna, the VSWR can be defined by the following formula:

$$\boldsymbol{VSWR = \frac{1+|\Gamma|}{1-|\Gamma|}} \qquad \textbf{(2.1.7)}$$

Γ : the reflection coefficient

If the voltage standing wave has a small value that means the value of the power transmission will deliver to the antenna better but does not necessarily mean the power delivered sometimes also radiated. The value of the ideal VSWR is 1 and that's will happen if the voltage of the transmission wave matched in this situation we will not have a return losses on the other hand if the voltage in the transmission waves haven't matched with the antenna that's mean there is some losses.

VSWR is a measure of how much power is delivered by a voltage source to an antenna. Which means that's an antenna dose not radiate all the received power; there is always amount of reflection voltage due to mismatching network.

Based on the VSWR concept we find that an antenna with VSWR less than 2 that's good enough to have a good impedance matching, on the other hand the transmission of more power to the radio can be lead to damage the radio.

2.2 Antenna types

Dipole Antenna

One of the most common antennas today in radio communications is a dipole antenna due to its simplicity. Dipole antenna has two rods or wires (poles) which are identical conductive elements, and fed by a signal source or feed energy. As shown in figure 2.2.1 below the two elements, current, and the voltage causes electromagnetic wave.

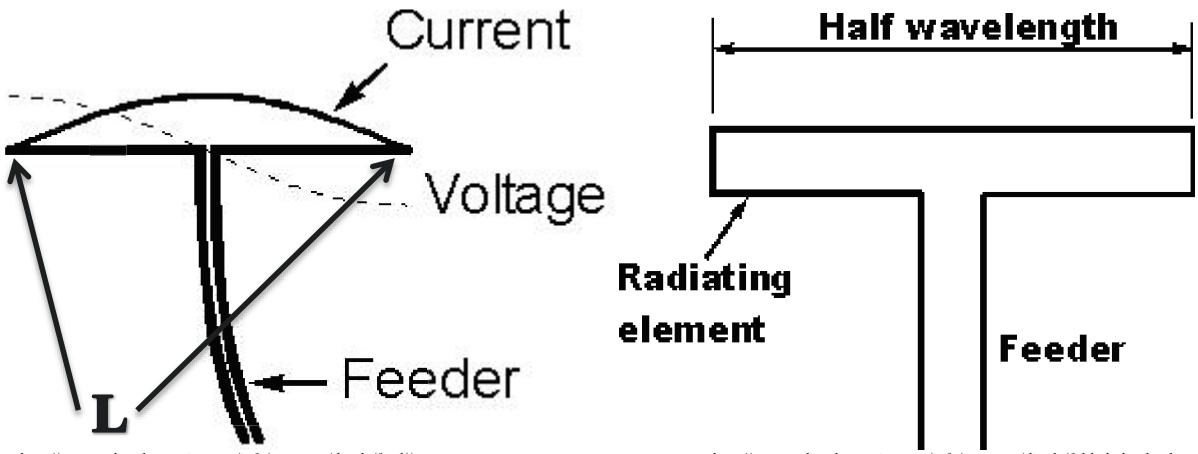


Figure 2.2.1 The half wave dipole antenna with center feed point, and to the right Half-wave folded dipole antenna.

Dipole antenna types

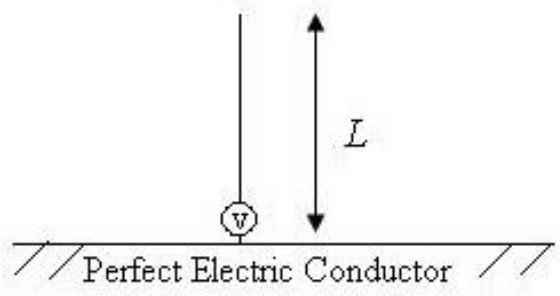
Half wave dipole antenna is the most popular type. But still there several types which could be used.

- **Short dipole**
  A dipole antenna is called a short dipole antenna if its rods or wires length (L) shorter than half of a wavelength (λ), for example if L< λ/10.
- **Half wave dipole antenna**
  As shown in figure # above a half wave dipole antenna consists of two wires each of them quarter of the wavelength, and the feeder is connecting to both wires.
- **Multiple half wave dipole antenna**
  A dipole antenna with odd multiple of half wavelength long.

- **Folded dipole antenna**
  A folded dipole antenna is half wavelength dipole but its wires are folded back as shown in figure 2.2.1.

Monopole antenna

Unlikely dipole antenna Monopole antenna is an antenna which consists of a one straight rod conductor, and is installed over a ground plane (conductive surface) as seen in figure 2.2.2. Monopole antennas are fed at lower end, near to ground plane which works as a reflector. In case the ground plane conductive and valid in size, the efficiency of the ground plane is as well as a vertically installed dipole.



http://www.antenna-theory.com/antennas/monopole.php

Figure 2.2.2 Monopole antenna with length L.

There are other common types of monopole antenna such as Whip, helical, T-antenna, and mast radiator. Monopole antenna like dipole antenna both have omnidirectional radiation pattern, for that the monopole it could be used in cell phones and other applications, for example indoor applications such as airplane, or shopping center.

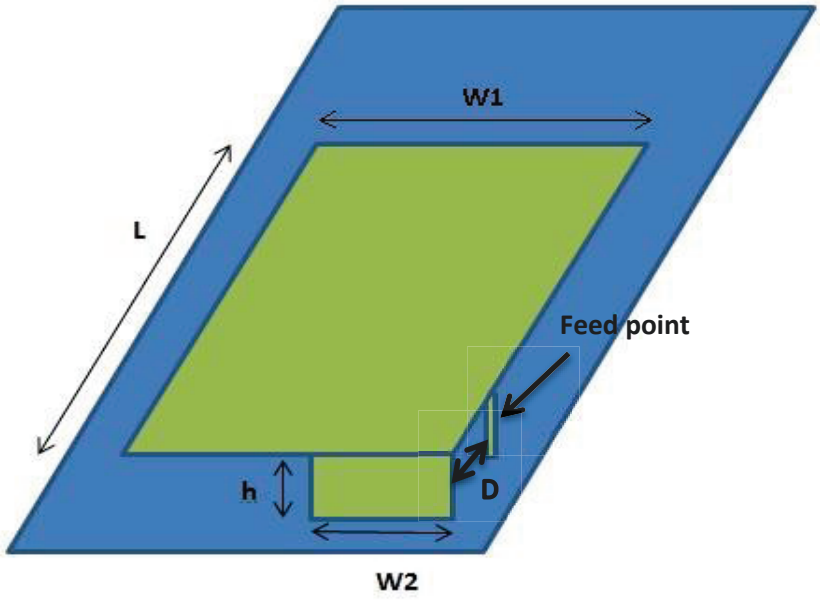
Monopole antennas are widely used in mobile communications, specially the quarter wave monopole -the length of the monopole L is approximately quarter of wavelength- which is a suitable for MIMO wireless communications systems, due to its low angle radiation and minimized ground losses [3].

For better efficiency of monopole antennas, many methods are applied, one of them type of the ground plane; The ground plane can be infinite or finite with different geometric shape such as spherical, cylindrical, or rectangular sheet. That's affect characteristics monopole specially radiation pattern, and less effects in monopole's impedance [4].

Based on fact that, monopole antennas are half the size (half length) of their dipole. So we can describe the directivity of monopole in term of its dipole, for example if the directivity of a dipole of length 2L has a directivity D1 [decibels], then the directivity of monopole with length L will be D1+3 [decibels], this because there is no radiation happens below the ground plane [4].

PIFA antenna

PIFA is the short name of Planar inverted-F antennas which is the most common type of cellular phone antenna nowadays, it has the same structure of microstrip antenna but it shorted with the ground because of this short pin the PIFA produce a quarter wave length. The main reason of name PIFA because has change of using wire to use ground plane. The PIFA structure consist of a rectangular planar which is installed on the ground plane, it is always parallel to reduce the height. Figure 2.2.3 is showing the geometric dimensions of the PIFA antenna the length which is L, The width which symbolized by W1, The short pin is W2, for height is h and the distance between the short pin and the feed point is D.



http://www.raymaps.com/index.php/planar-inverted-f-antenna-pifa/

Figure 2.2.3 PIFA antenna and its geometric dimensions.

The structure of the PIFA can deal with different frequency band

- Single band
- Multiband
- Reconfigurable

The main advantages of using PIFA antenna

1- Due to the small size of PIFA antenna it gives the availability to insert the antenna inside the cellular phone.
2- PIFA has availability to get high gain for both polarization states vertically and horizontally there for it can receive the reflection wave easily from different directions.
3- The backward radiation of the PIFA has be reduced that's mean the electromagnetic waves power has reduced which lead us to get a good deal to less damage of the human health.
4- The design and the material of the PIFA are not costly on top of that it has a very high efficiency also It is easy to fabricate.

As we know the development of communication always come up with a new evaluation but until now there are no antennas can work without disadvantages, So there is a problem in PIFA antenna could not be solved until now which is the sensitivity of the bandwidth but in general PIFA prove that is the best cellar phone antenna have been used regarding to the comparing between the advantage and the disadvantage of the antenna.

There are a few di-electrical material can be used in PIFA, and when the air is the one which is used, the gain will be better comparing with other material.

Fractal antenna

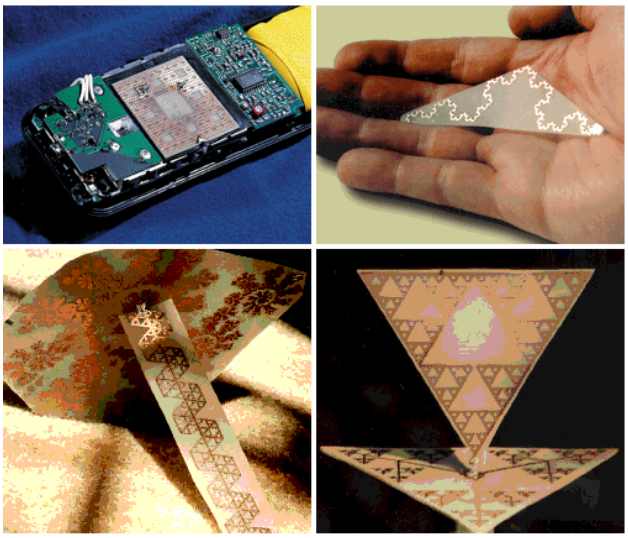


Photo was loaded from http://www.antenna-theory.com/antennas/fractal.php

Figure 2.2.4 Triangle Fractal Geometry. Upper Left: Patch Antenna Fractal. Lower Left: Monopole Fractal. Lower Right: Bowtie Monopole Fractal Antenna.

One of the problems of popular designs of antenna, that's the antenna which is smaller than a quarter of the wavelength, is very sensitive to a narrow band of frequencies. This problem is a significant issue for mobile phone, where the requirements are always to minimize the size of the mobile phone which leads to the problem of sensitive narrow band frequencies.

As shown in figure 2.2.4 the fractal antenna has a complex shape which offers many electric current modes to exist, that's causes, the increasing in radiation, consequently very wide bandwidth.

Fractal antenna designs can solve sensitive narrow band frequencies problem. Experiments have shown that antennas built with only a small number of iterations of a fractal process can exhibit sensitivity at several frequencies, and it is worthy to mention that's fractal antenna can work efficiently more at one quarter of the wavelength than other antennas types [5].

Geometry of fractals

The shape of a fractal can be formed by iterative mathematical process called Iterative Function Systems (IFS) [6] and by using affine transformations as following:

W(x, y) = (ax+by+e, cx+dy+f)

a,b,c,d control rotation and scaling e , f control linear translation by repeating the upper calculations, we get the IFS sequence that converges to the final image.

Chapter 3

Mobile phones antenna design

Mobile phones consist of several antennas for several purposes such as Wi-Fi antenna, GPS antenna, low and high frequencies antennas. Each type of antennas is designed according to certain considerations, of course, the small sizes the main issue.

As it is mentioned in the previous sections of this thesis. Dipole antenna is a useful type of antennas; easy to be constructed, and integrated in cell phones

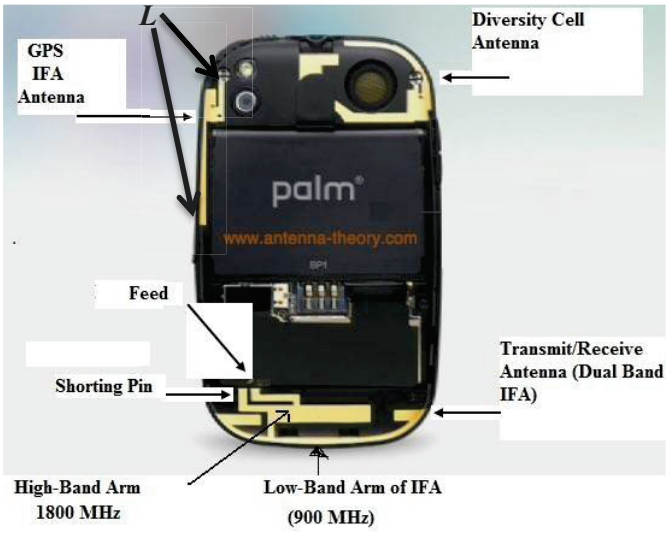
3.1 Mobile phone GPS antenna design

GPS antenna is an omnidirectional small antenna with only receive mode that can be one of main design's considerations. The GPS frequency is 1.575 GHz, with practically no bandwidth, and the GPS satellites are RHCP (table 3.1.1) to avoid Faraday rotation problems, and there is one other benefit, that's circular polarization does not need rotational alignment of a circularly polarized antenna at mobile phone.

For a maximum received power the user terminal's antenna must be RHCP. User terminal's antenna which is linearly polarized will have a loss 3dB in received power due to polarization issue. By looking at the power density of the GPS received wave. We find that it is extremely low level which equals to -160dB. Therefore, we need an efficient antenna at the receiver.

Table 3.1.1 Cell phone GPS antenna specifications and goals [4].

Transmit and receive frequency	Receive only 1.575 GHz
Efficiency	-3 dB to -9 dB
Polarization	RHCP
Quarter wavelength(L)	About 4.75cm or 1.87 inches



http://www.antenna-theory.com/antennas/aperture/ifa.php

Figure 3.1.1 Palm Pre Antennas viewable by Removing Back Cover.

As shown in figure 3.1.1 in this design, quarter wavelength IFA antenna is used. The ground plane (the chassis) of the mobile phone is considered as one arm of dipole antenna and the feeder is placed on opposite side at a point which helps for matching impedance. The GPS antenna is placed on the top of the mobile phone to guarantee best performance; that's because the users hold the mobile phone vertically when they want check current location for example, and in that case their hands will be on lower part of mobile phone, far from GPS antenna.

As obvious in figure 3.1.1 IFA antenna is designed with its geometric dimensions to get the better efficiency, and of course to take into account the available low profile for the antenna in mobile phone.

To obtain circularly polarized dipole antenna, we use two crossed dipoles to provide the two orthogonal field components, and if the two dipoles are fed with a 90° time-phase difference the polarization will be circular along zenith [28].

Single band PIFA Antenna design

GPS antennas are received mode antennas. So we do not have to worry about SAR measurements. In this section we will design a single band antenna for GPS use and simulate the model by COMSOL.

Model descriptions

The desired bandwidth is a single frequency 1.575 GHz (table 3.1.1). Thus we try to make the resonant frequency of the PIFA antenna close to 1.575 GHz to provide high receiving efficiency as much as possible. The gain of the antenna should be between -3dB to 0dB.

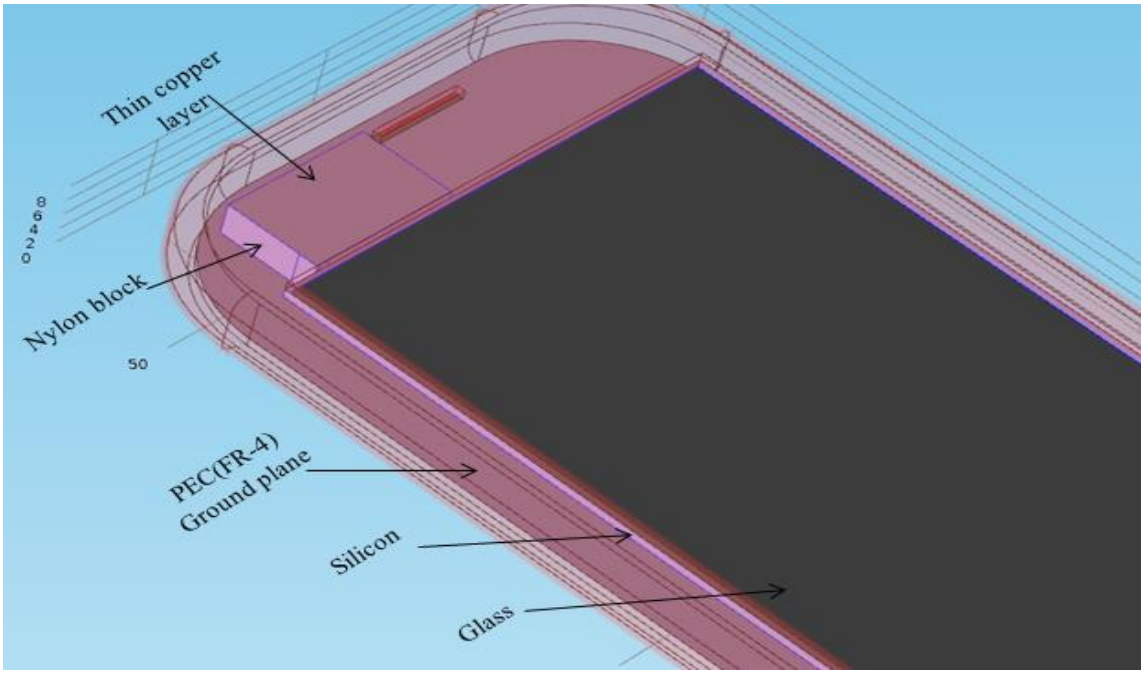


Figure 3.1.2 Zoom view of PIFA antenna block. It consists of Nylon block and the radiating part and the materials which are used in design.

Calculations

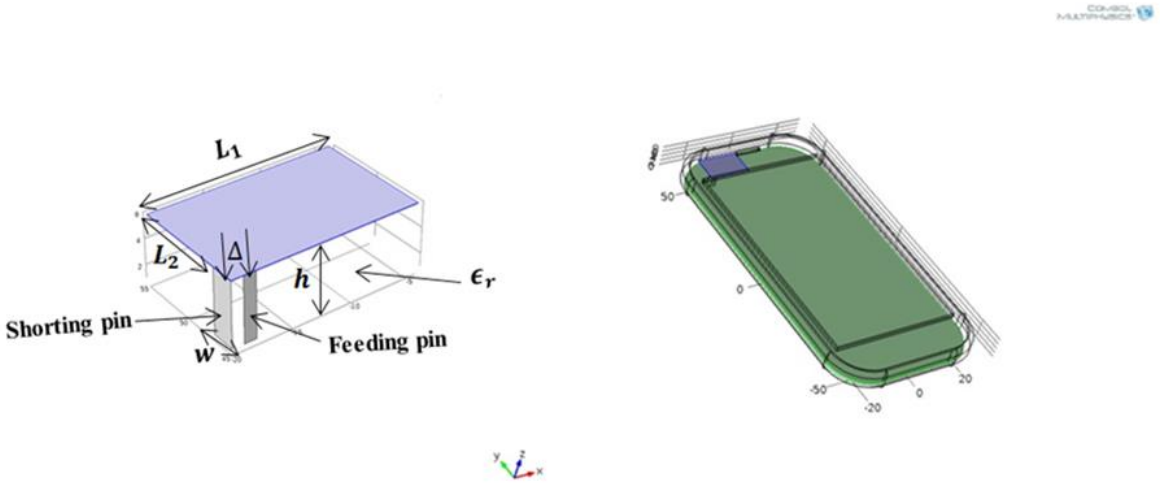


Figure 3.1.3 The antenna to the left and its dimensions $L_1 = 20mm, L_2 = 10mm, w = 2mm, h = 4mm$ $\epsilon_r = 3.8$, $\Delta = 0.2mm$ *to the right the casing's dimensions have length =119mm, width =60mm.*

The resonant wavelength of a PIFA antenna can be calculated as following:

$$\boldsymbol{L_1 + L_2 - W = \frac{\lambda_0}{4}} \qquad \textbf{(3.1.1)}$$

And the relation between the resonant wavelength and the resonant frequency can be determined by the equation:

$$\boldsymbol{\lambda_0 = \frac{c_0}{f_0\sqrt{\epsilon_r}}} \qquad \textbf{(3.1.2)}$$

λ_0: Resonant wavelength of PIFA.

c_0: The speed of light in space.

f_0: Resonant frequency.

ϵ_r: Relative permittivity.

Mathematically, PIFA antenna's size can be reduced by using factor $\sqrt{\epsilon_r}$ (equation 3.1.2)

Therefore if we use a material with higher ϵ_r ,that reduces the size of the PIFA which is a useful way to get a low profile antenna but that could affect the gain of the antenna as well. It is a big challenge to keep the efficiency without any effects when the small size is required, so the PIFA dimensions which are used in this design were chosen after many simulations, and based on the best efficiency, lowest value in return loss S11, and the smallest size. We determined the dimensions.

By applying equation 3.1.1 we find that:

$20 + 10 - 2 = \frac{\lambda_0}{4}$, $\lambda_0 = 112$ mm

Hence,

$$f_0 = \frac{3 \times 10^8}{112 \times 10^{-3}\sqrt{3.8}} = 1.374 \text{ GHz}$$

The calculated resonant frequency is not very close too much to the desired frequency 1.575 GHz. But by increasing and decreasing the gap impedance change, the resonant frequency will shift to upper or lower than the calculated resonant frequency. Increasing and decreasing of the impedance gap affects the input impedance, consequently effects at VSWR for the resonant frequency.

Simulations

In simulation, first we installed the antenna to left of the casing figure 3.1.3 and inserted the materials as shown in figure 3.1.2 and with dielectric constant as table 3.1.2 in addition the outer shell of the casing is simulated with Polytetrafluoroethylene (PTFE). We use whole chassis as a ground plane for better efficiency.

A 50 Ω lumped port is used to excite the antenna and determine the input impedance. The lumped port is mounted between two metallic boundaries: the vertical feeding and the ground plane of FR4 board as seen in figure 3.1.2. The distance Δ, the impedance matching gap affects significantly the matching impedance. So another strip shorted to the ground plane is added. The dielectrically material which will be used for the antenna block is Nylon.

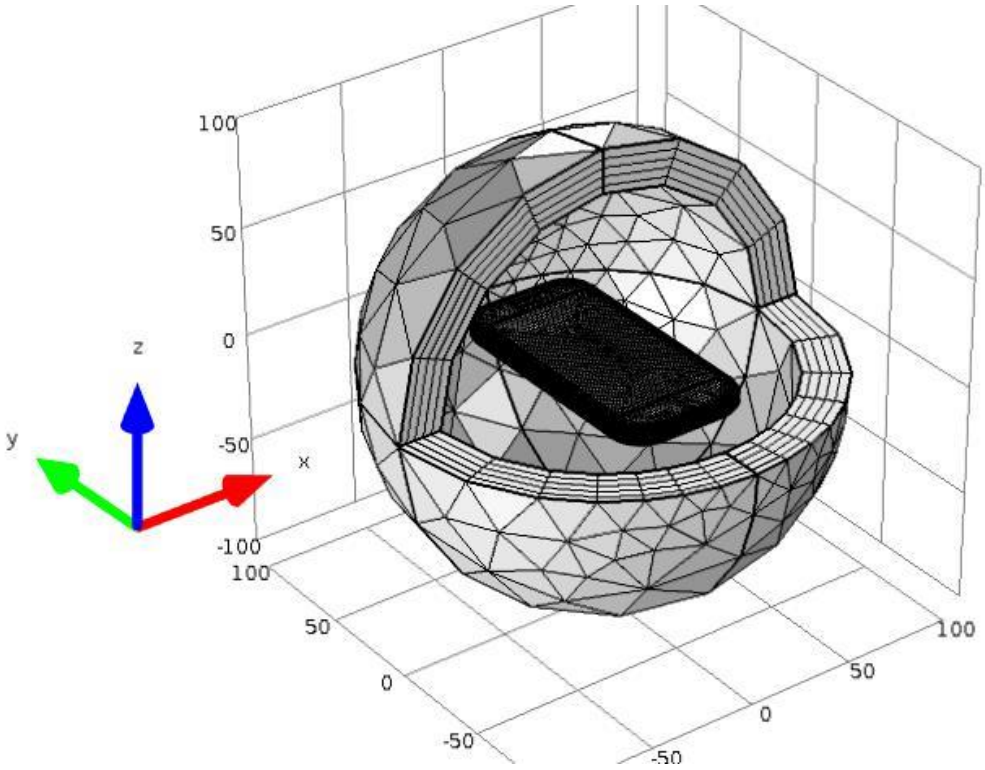


Figure 3.1.4 Mesh model for the whole simulated model: PIFA antenna, casing of the mobile phone and the surrounded perfect matched layers.

The model consists of several domains; every domain has a certain material figure 3.1.2. In addition there is one more domain encloses the casing. This domain is simulated as a sphere with radius 100mm, material air, and has five perfect matched layers in order for the radiation to be able to travel anywhere as shown in figure 3.1.2. The metal part of the antenna element at frequency 1.575 GHz can be modeled using perfect electric conductor boundaries.

Table 3.1.2 the materials of the simulated model and its Dielectric constant according COMSOL's materials libraries except Nylon.

Material	Dielectric constant (ϵ_r)
Air	1
FR-4	4.5
Nylon	3.8*
Glass (quartz)	4.2
Silicon	11.7
PTEF	2.1

* The Dielectric constant of Nylon accordinghttp://www.professionalplastics.com/professionalplastics/ElectricalPropertiesofPlastics.pdf

Results and Analysis

E-field norm on the xy-plane slice is shown below in figure 3.1.5. The plot shows that the electric field is strong at one of the top metallic surface shell far from the feeding point. This looks alike the E-field distribution of a quarter wavelength monopole antenna, which the PIFA derived from.

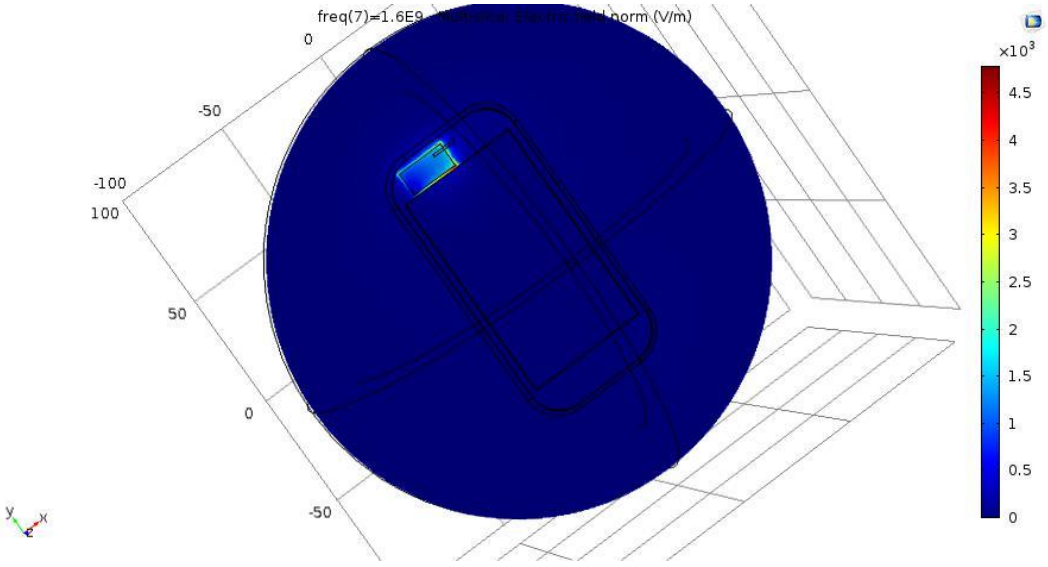


Figure 3.1.5 E-field norm distribution on the top of the PIFA.

The polar plot of the far field radiation pattern of the antenna is shown in figure 3.1.6 As obvious the antenna gain on xy-plane varies from about -6dBi to 1.5dBi. The azimuthal radiation pattern is not Omni-directional any more, since the antenna is mounted on the ground plane and miniaturized.

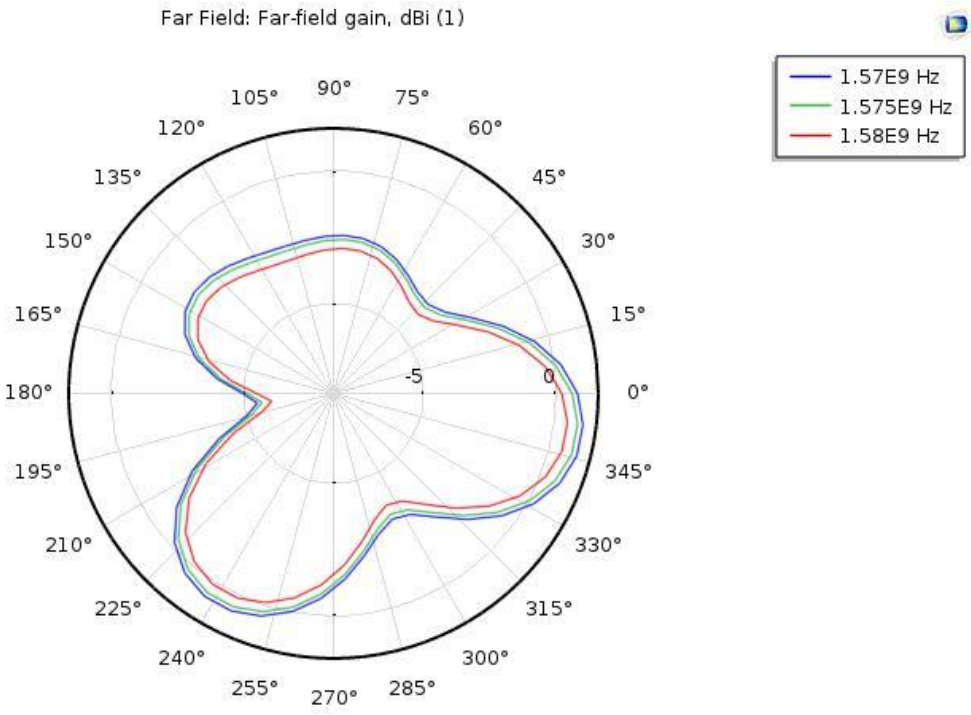


Figure 3.1.6 Antenna gain pattern on the xy-plane.

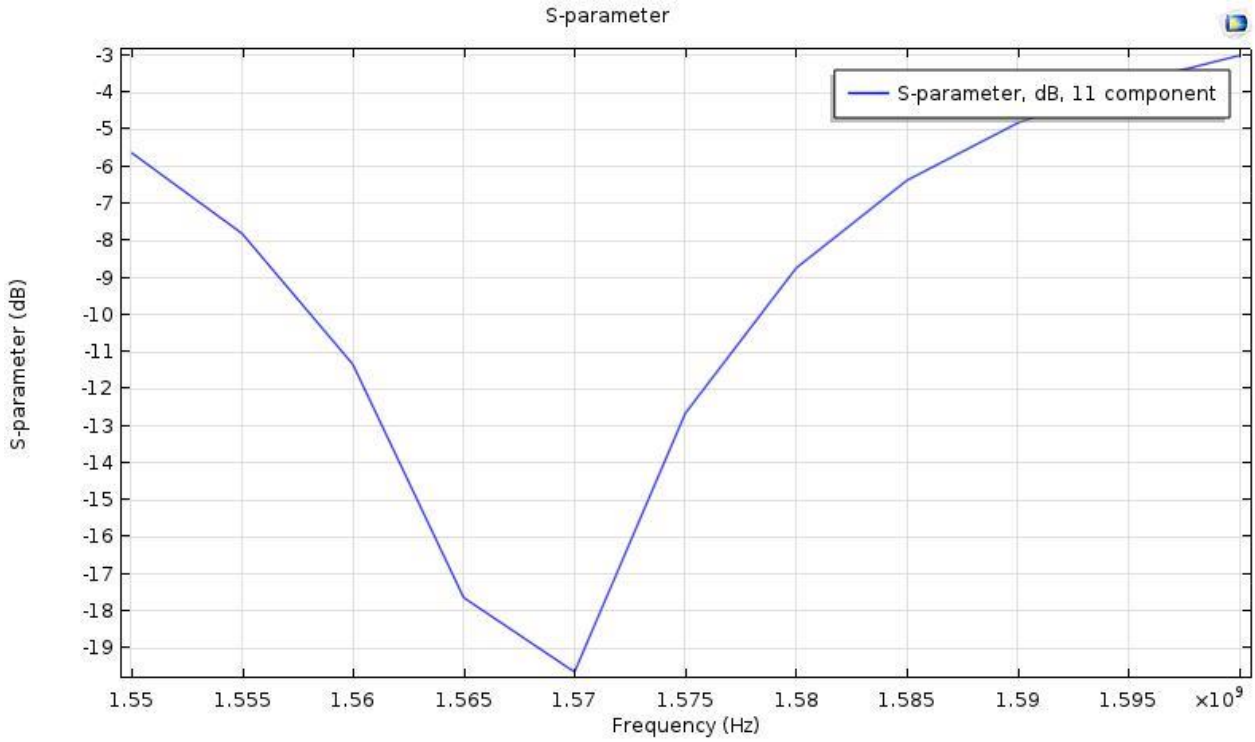


Figure 3.1.7 Return loss S-parameters (S11) for the antenna by simulation.

S-parameters (S11) measurements indicate that at 1.575GHz is -13dB which means that the reflected power is 5%. This describes how well the antenna input impedance is matched to the 50 Ω reference impedance.

Antenna's bandwidth regarding figure 3.1.7 is a narrow bandwidth:

$$Bandwidth \approx \frac{1.577 - 1.557}{1.570} \times 100 = 1.2\%$$

The wide bandwidth for GPS antenna is not required. So the bandwidth above is sufficient.

Keeping the feeding point near to the shorting pin as much as possible reduces the antenna size but that causes a narrow bandwidth. The electrical characteristics of PIFA are affected by the size of ground plane. By varying the size of ground plane, the efficiency and the bandwidth are changed; to increase the bandwidth we need the whole possible free space in the chassis. In simulation we use the whole chassis as ground plane in order to get a better efficiency.

The effective of the PIFA antenna geometric dimensions on the resonant frequency can be summarized by the following table 3.1.3

Table 3.1.3 Effect of geometric dimensions on resonant frequency.

	L_1	L_2	W	H
f_0 ↑	↓	↓	↑	↓

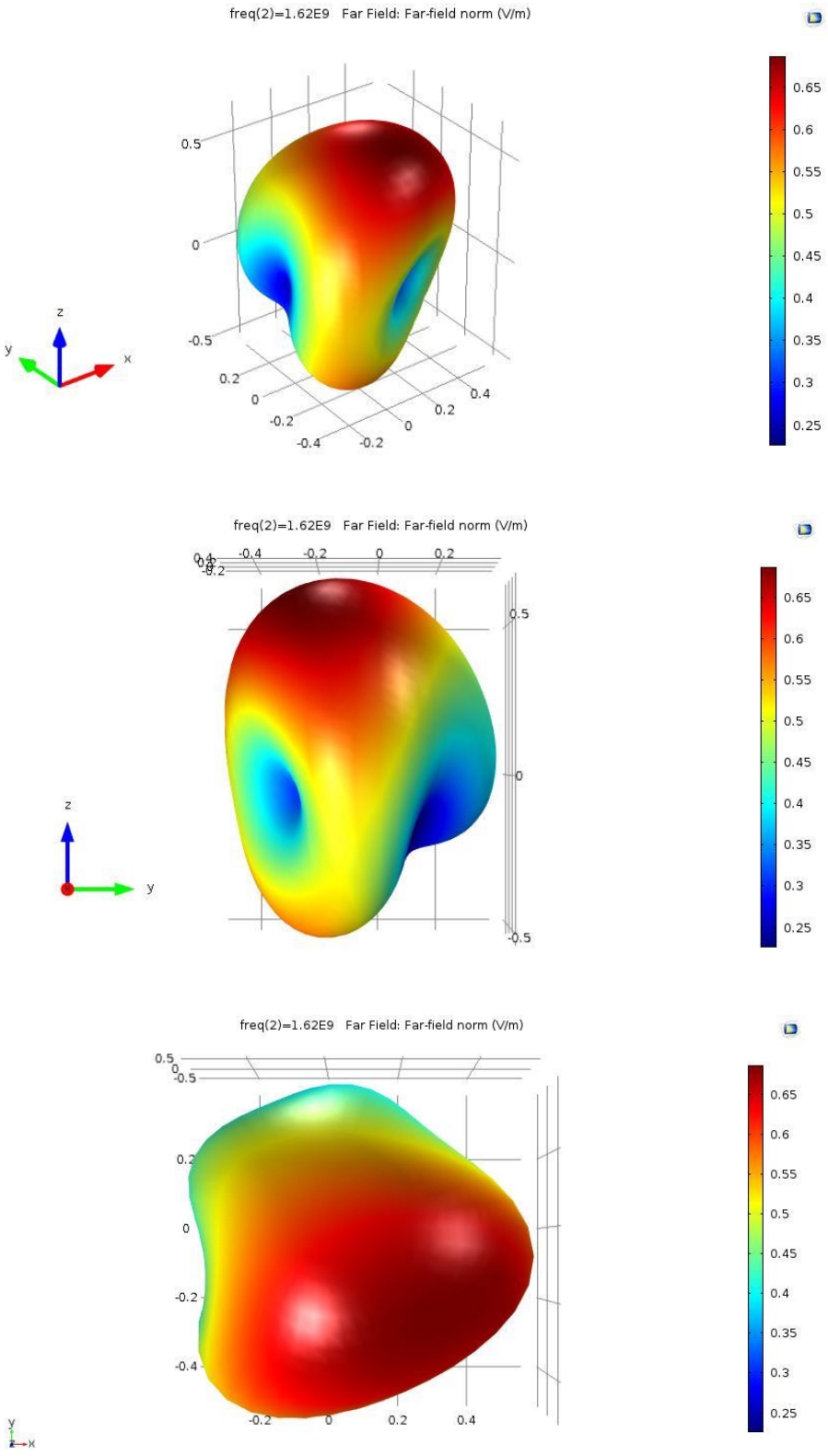


Figure 3.1.8 3D far-field radiation pattern shown from three different angles.

Table 3.1.4 several mobile phones models and its internal type antenna [29].

Model	Antenna	Length	Width	Height	Effective Volume (cc)	Form Factor	Frequency Bands
Apple iPhone 2G	Planar Monopole	55.6	22.8	10.4	6.9	Candybar	GSM850/900/1800/1900
Apple iPhone 3G	Planar Monopole	55	20	8.5	9.0	Candybar	GSM850/900/1800/1900+3G
Apple iPhone 4	Planar Monopole	58	14	6.1	4.8	Candybar	GSM850/900/1800/1900+3G
ASUS M307	Planar Monopole	35	24	0.2	1.6	Clamshell	GSM900/1800/1900
ASUS P525	PIFA	34	20	7.6	5.1	Candybar	GSM850/900/1800/1900
BenQ Siemens EF-71	Planar Monopole	38	10.1	5.3	2.0	Clamshell	GSM900/1800/1900
Blackberry 8100	Planar Monopole	42	9	6.7	7.7	Candybar	GSM850/900/1800/1900
Blu 233/Sendo M570	Planar Monopole	35	16.4	1.5	0.9	Clamshell	GSM900/1800
Geo GC688	PIFA	36.5	10.7	5.8	2.3	Slider	GSM850/900/1800/1900
Hagenuk	Slot	72.4	49.6	6.0	12.6	Candybar	GSM900
Motorola E398	PIFA	37.2	20.1	8.7	5.5	Candybar	GSM900/1800/1900
Motorola KRZR K1	Planar Monopole	37.5	7.6	7.0	2.4	Clamshell	GSM850/900/1800/1900
Motorola L2000/P7389	Helix	44.2	5	5.0	2.1	Candybar	GSM900/1800/1900
Motorola L6	Planar Monopole	36.1	7.5	7.0	1.9	Candybar	GSM900/1800/1900
Motorola T193	PIFA	27	17.8	8.1	2.9	Candybar	GSM1900
Motorola T720i	Helix	29	10	10	1.6	Clamshell	GSM900/1800
Motorola V690	PIFA	38	16.9	9.5	6.1	Clamshell	GSM900/1800/1900
Motorola W208	PIFA	35.6	24	6.5	2.1	Candybar	GSM900/1800
Nokia 2626	PIFA	38	22.3	7.7	3.0	Candybar	GSM900/1800
Nokia 2652	PIFA	41.4	18.6	6.6	5.1	Clamshell	GSM900/1800
Nokia 5210	PIFA	34	18	6.6	4.4	Candybar	GSM900/1800
Nokia 5300	Planar Monopole	36.5	8.4	9.3	2.8	Slider	GSM900/1800/1900
Nokia 5320	PIFA	42.6	28.5	8	6.4	Candybar	GSM850/900/1800/1900+3G
Nokia 5500	PIFA	34.4	38.8	8.5	10.5	Candybar	GSM900/1800/1900
Nokia 6030	PIFA	38	22.3	9.1	6.6	Candybar	GSM900/1800
Nokia 6100	PIFA	37.9	28.6	6.3	6.1	Candybar	GSM900/1800/1900
Nokia 6108	PIFA	37.9	28.6	5.6	5.5	Candybar/ Flip	GSM900/1800/1900
Nokia 6111	Planar Monopole	41.4	9.1	6.8	2.5	Slider	GSM900/1800/1900
Nokia 6210	PIFA	42.6	19.8	9.9	8.4	Candybar	GSM900/1800

3.2 Wideband PIFA Antenna Design

PIFA antenna is used widely in mobile phones today as shown in the table 3.1.3, that's due to many advantages of PIFA which are mentioned in chapter 2 Antennas types. Hence in this section PIFA antenna will be used to design a wideband antenna and simulate the model in COMSOL.

Model descriptions

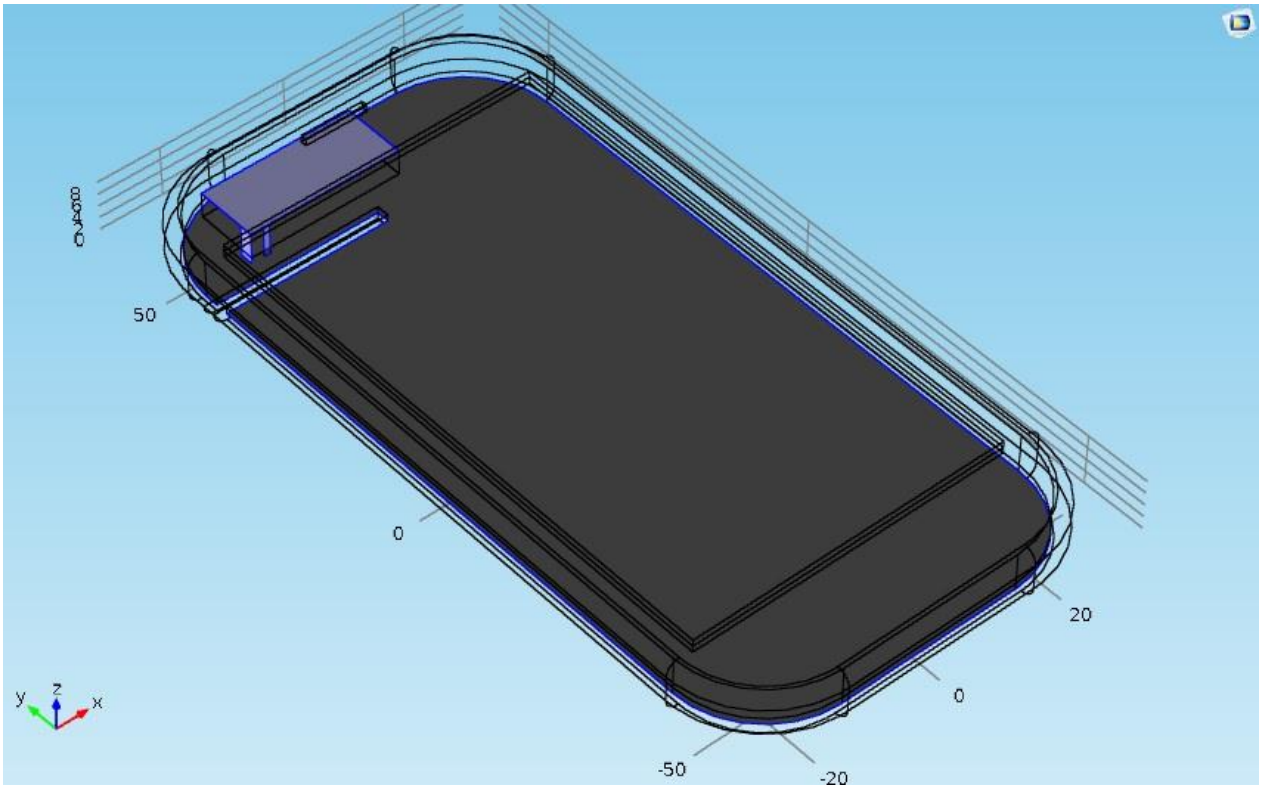


Figure 3.2.1 the wideband PIFA antenna is mounted to left top of the slotted ground plane.

PIFA antenna design using slot technique, with desired range of frequency from 1800MHz to 2600MHz. This important range cover GSM (1800MHz & 1900MHz), UMTS (2100MHz), Bluetooth and Wi-Fi (2.4GHz), and LTE system (2.3GHz, 2.5GHz, and 2.6GHz).

The wideband PIFA antenna design using slot technique model consists of the same materials which are used in previous design table 3.1.2, except the dielectric material between the PIFA and the ground plane which is air with dielectric constant (ϵ_r) equals to 1 to have a better efficiency and a wider bandwidth.

Calculations

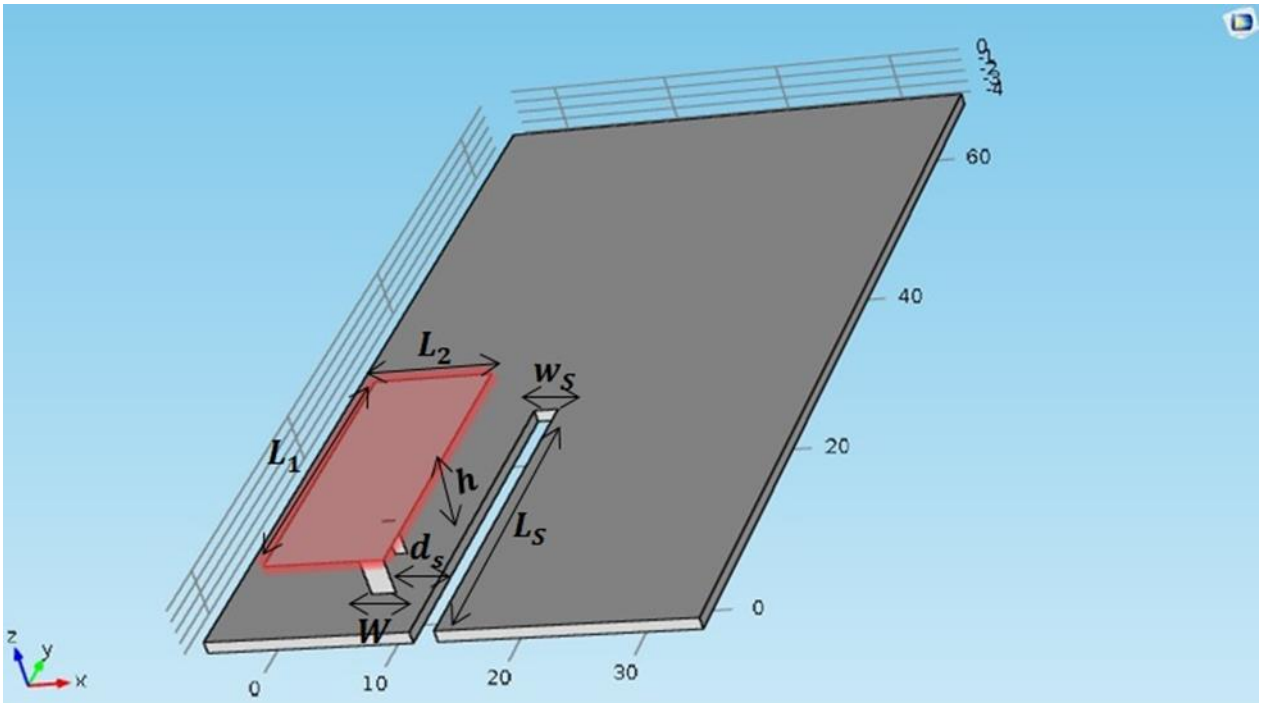


Figure 3.2.2 The illustration dimensions of the PIFA $L_1 = 24mm$, $L_2 = 10mm$, $h = 4mm$, $W = 2mm$ *and the ground plane with its slot dimension* $d_s = 5m, Ls = 28mm$, $W_s = 2mm$.

The resonant frequency of the PIFA antenna can be calculated using 3.1.1 and 3.1.2 formulas respectively.

$24 + 10 - 2 = \frac{\lambda_0}{4}$, $\lambda_0 =$ 128mm

$f_0 = \frac{3\times10^8}{128\times10^{-3}\times\sqrt{1}} = 2.343GHz$

The desired bandwidth percentage:

$f_{center} = \frac{2600+1800}{2} = 2200MHz$

$Bandwidth\% = \frac{2600-1800}{2200} \times 100 = 36.36\%$

If the bandwidth percentage higher than 20%, that's bandwidth is considered as a wide bandwidth.

Simulations

In this design we use a slotted ground plane in order to improve PIFA bandwidth. A slot in a ground plane creates different paths of current flow which leads to kind of diversity in electric field, consequently a wider bandwidth. The length of the slot should be proportional with the desired resonant frequency. The simulations are performed for two cases, one without the slot and other with slot, for different positions of the slot on the ground plane and for different dimensions of PIFA except the width of the planar which is fixed at 10mm and width of the short pin is 2mm.

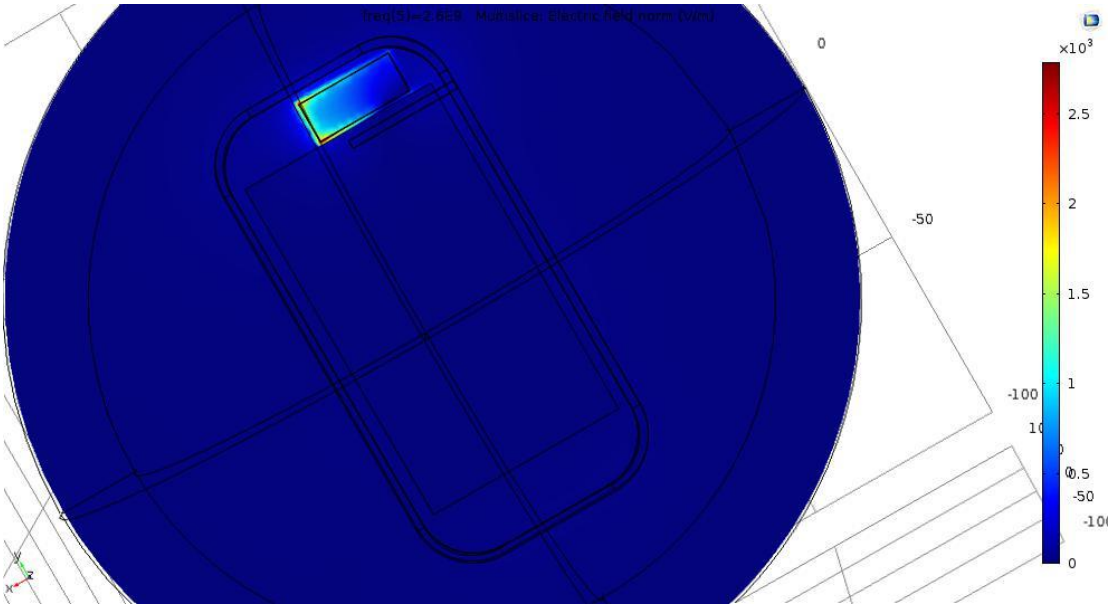


Figure 3.2.3 Back view of E-field norm distribution on the top of the wideband PIFA.

In figure 3.2.3 the plot of the E-field shows that the electric field is strong at one of the top metallic surface shell far from the feeding point on both corners of the planar which similar to E-field plot in previous design.

Figure 3.2.4 shows the far-field gain in dBi for every frequency center of the desired bands and in figure 3.2.5 the 3D far-field radiation pattern of wideband PIFA shown from three different angles.

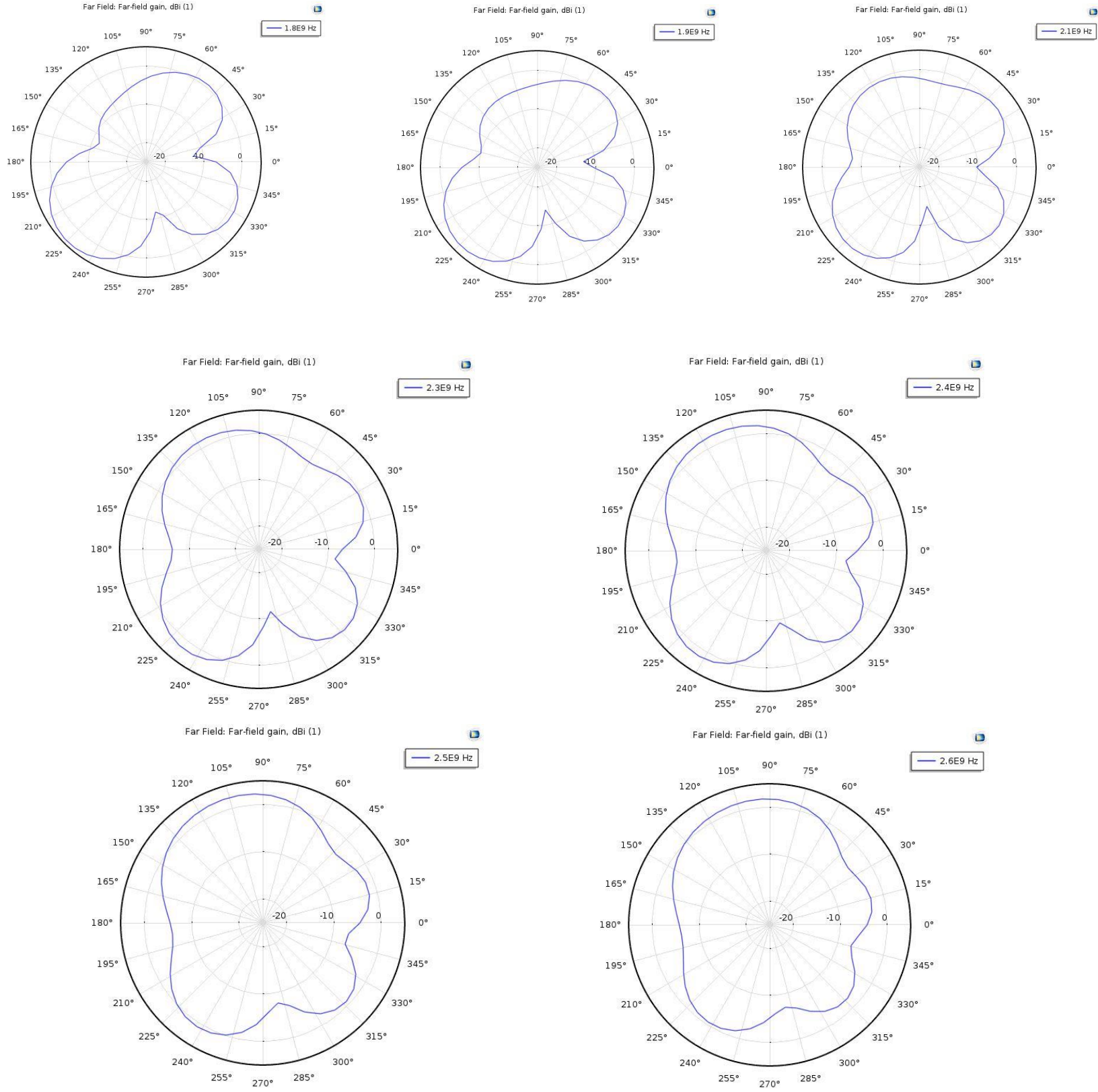


Figure 3.2.4 Antenna gain pattern on the xy-plane for each frequency band.

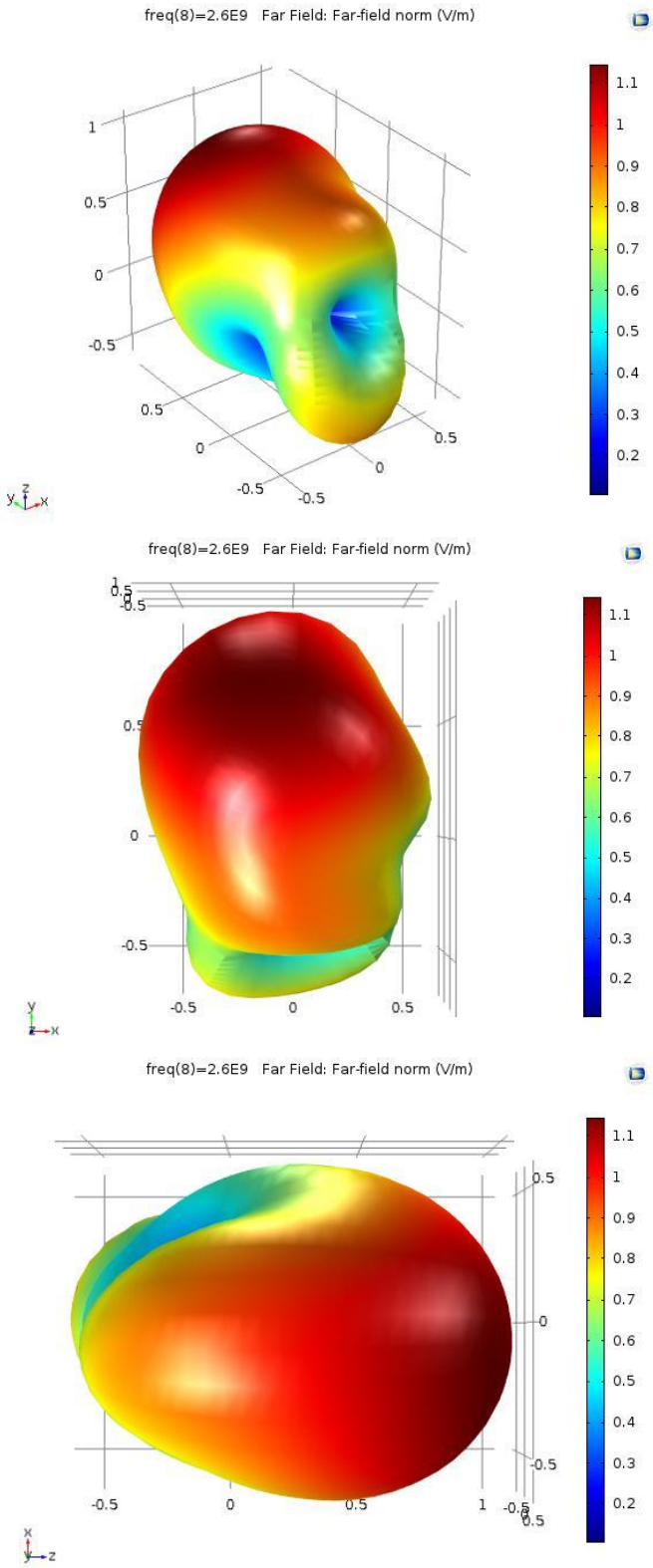
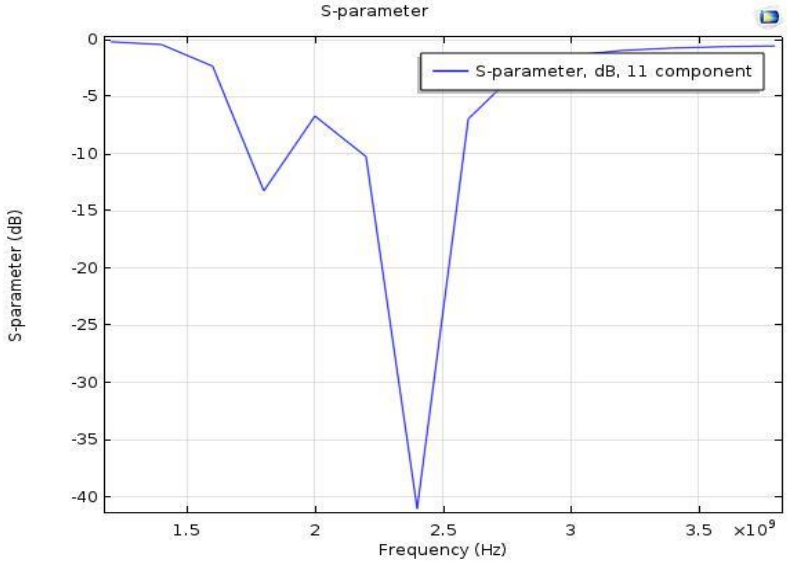


Figure 3.2.5 3D far-field radiation pattern of wideband PIFA shown from three different angles.

Results and Analysis



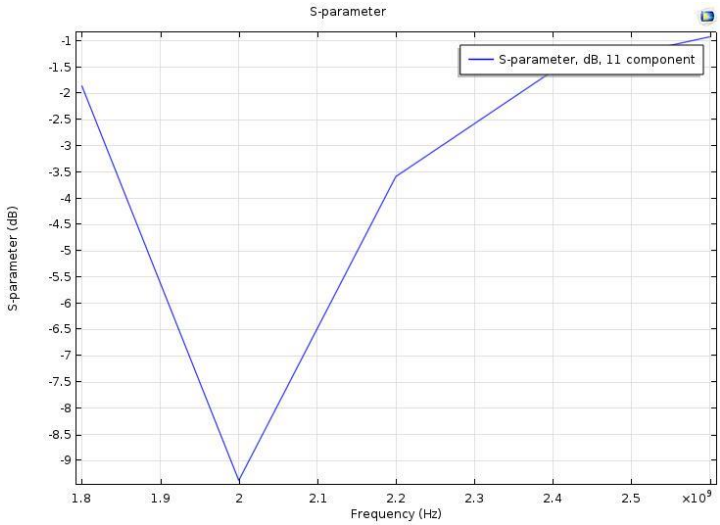


Figure 3.2.5 (a) Return loss (S11) for the wideband PIFA using slotted ground plane by simulation.

(b) Return loss (S11) for PIFA without slot.

As shown in figure 3.2.5 (a) the resonant frequency is 2.4GHz with -40dB and return loss of the frequencies 1800MHz,1900MHz,2300MHz,2400MHz and 2500MHz is $\leq -10dB$. At 2600MHz is -8dB and at 2100MHz is -9dB which it's still reasonable. With considering 2600MHz has a sufficient low return loss, the bandwidth will be 36.36%.

Figure 3.2.5 (b) shows the PIFA antenna without using a slot on ground plane with resonant frequency at 2GHz, return loss -10dB and the bandwidth can be calculated as following:

$$Bandwidth\% \approx \frac{2.01-1.99}{2} = 1\%$$

Based on what is mentioned above, we can conclude that's the slot has a big significant effect on either bandwidth or on better matching impedance network. In bandwidth there is a huge difference; without using slot the bandwidth is 1% and with slot is 36.36%. Even if we exclude the band 2600MHz the bandwidth 32.36% which is still much wider rather, comparison is nonessential. On other hand the effect of the slot on return loss (S11) is clear as well. Where PIFA with slotted ground plane resonate at 2.4GHz with -40dB, and without slot resonate at 2GHz with poor return loss -10dB.

As seen in figure 3.2.4 it's obvious that all frequencies bands have an Omni-directional behavior with reasonable gain table 3.2.1 summary the maximum and minimum gain.

Table 3.2.1 Simulated gain of the frequencies which shown in figure 3.2.4.

Frequency	Maximum gain	Minimum gain
1800MHz	4.95dB	-12.48dB
1900MHz	3.97dB	-13.86dB
2.1GHz	2.89dB	-14.72dB
2.3GHz	2.54dB	-11.31dB
2.4GHz	2.78dB	-9.33dB
2.5GHz	2.89dB	-7.79dB
2.6GHz	2.25dB	-7.1dB

In this design several methods are followed in order to improve the Bandwidth and the gain, where the relation between the Bandwidth and the gain is not linear. Using dielectric material with high dielectric constant causes a degraded gain; on other hand the effects on Bandwidth could be neglected.

Techniques which are used to increase the Bandwidth for proposed PIFA:

- Bandwidth depends very much on the size of the ground plane. So for better performance, the whole available size of the chassis should be used as a ground plane. Reducing the ground plane can effectively limit the bandwidth of the PIFA antenna.
- Using slotted ground plane: using a slot with proper length to get other resonant frequencies.
- Using Air as a dielectric material between the PIFA element and the ground plane this technique improves the Bandwidth and enhances the gain.

A wideband PIFA antenna has been designed and presented. The proposed PIFA antenna occupies a compact envelope dimension of $24 \times 10 \times 4mm^3$ while covering the required wide band with a sufficient impedance matching (S11 ≤ -10 dB) covering GSM (1800MHz &1900MHz), Bluetooth & Wi-Fi (2.4GHz), and LTE system (2.3GHz, 2.5GHz) except UMTS (2.1GHz) and LTE (2.6GHz) which has acceptable return loses.

3.3 Wi-Fi and Bluetooth Antenna

Wi-Fi and Bluetooth antennas work by transmitting and receiving the electromagnetic waves from the antenna to the receiver. They are one of the most useful services in the new smart phones, Wi-Fi antenna is the smallest antenna could be finding in the mobile phones because it works with the highest frequencies.

One of the best advantages of the Wi-Fi antenna that can support dual-band, the first band is from 2400 MHz to 2484MHz and also can support 5150-5850MHz; therefore the Wi-Fi antenna can cover the band of the Bluetooth too.

Nowadays, the most useful Wi-Fi antenna in mobile phone that's is connect to small chip which work for the Wi-Fi and the Bluetooth antennas in the same time, also we can have two antennas one for the Wi-Fi antenna which cover 5GHz and the other one for the Bluetooth which cover 2.4GHz but it is not common to use two types of antennas it will need more space and it will be costly than use one antenna.

As we mentioned Wi-Fi antenna is the smallest antenna in the mobile phones, the antenna half wavelength which covering the Bluetooth 2.4 GHz supposed to be 6.25 cm and at 5GHz which is mostly dipole antenna the half wavelength for sure will be smaller, it is 3 cm. also the quarter wavelength could be used which mean that the size of the antenna will be much smaller which is usually PIFA or PMA.

The quality of the Wi-Fi and Bluetooth antenna connection depend mainly on the gain of the antenna as we have mentioned in the second chapter, by determining the antenna power gain we can identify the efficiency and the directivity of the antenna, the order of the Wi-Fi mobile phone antenna efficiency from -6 dB to -2dB.

In case the primary antenna of the mobile phone does not cover the Bluetooth and Wi-Fi frequencies the Wi-Fi and Bluetooth antenna integrated on the top of the phone close to the GPS antenna due to the hand held of the user.

Chapter 4

Specification Absorption Rate (SAR)

As the world care about the development of wireless communication and with the really huge goals they have achieved on the other hand we should consider the human health and the risk which is effect negatively in the human health.

The meaning of SAR is a short name of specification absorption rate which is the measurement of the energy has absorbed by the human body during transmit the radio frequency electromagnetic field. The human body absorbed the energy that's mean we will lose some energy the second problem that's mean will affect the human body badly.

SAR can be calculated by integrating or averaging over 1 gram or 10 gram:

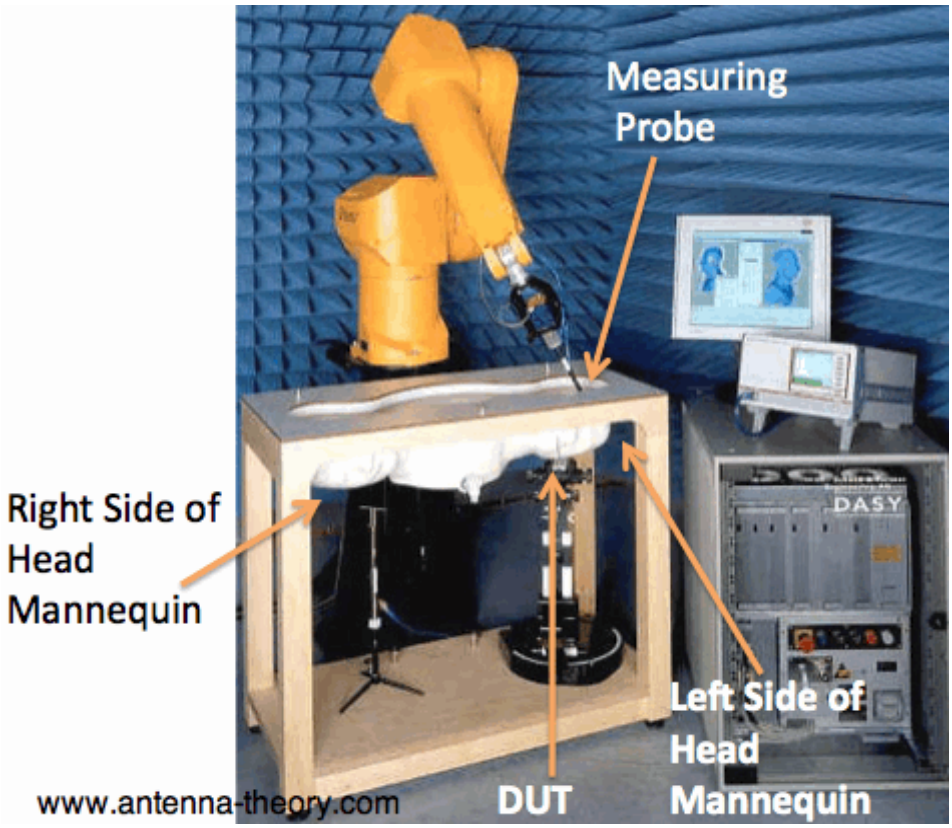
$$\boldsymbol{SAR = \int \frac{\sigma(r)|E(r)|^2}{\rho(r)} dr} \qquad \textbf{(4.1)}$$

Regarding to the equation 4.1 it explains that SAR is a function of the induced electrical field which is from the radiated energy E it can be measured by $volts\ /\ meter$, the electrical conductivity σ which can measured in $Siemens\ /\ meter$ and the mass density ρ can be measured in $g\ /\ (cubic\ meter)$, finally the unit of the SAR is $W\ /\ kg$.

The SAR have different values regarding to the design of the mobile phone and the location of the antenna in the mobile phone, therefore always, the mobile phone antenna located on the bottom of the phone to keep the radiate as far as much of the user. Low SAR means that it is safer than the high SAR while all the mobile phones has radio frequencies. On the other hand we don't forget to mention that the SAR will effect on the quality of the power by reducing the power level because of the absorption.

4.1 How To Measure The SAR?

Measuring the SAR by the DASY measurement system as shown in figure 4.1.1



http://www.antenna-theory.com/definitions/sar.php

Figure 4.1.1 The DASY SAR measurement system.

If we look at the figure 4.1.1 it shows different equipment which gives us availability for picturing the same situation of using the mobile phone in the real life.

The hollow tub in the figure 4.1.1 has the same shape of the right side and the left side of the human head, also there is a yellow robot with long arm which has in the end of the arm a measuring probe, the measuring probe moves around the all the region of the right side of human head and the left side too to lead us the measuring the SAR. But to get the correct measuring of the SAR we should have the correct conductivity and correct density therefore the tub must be filled with special fluid which has almost the same properties of the human tissue.

It is really worth to mention that the fluid has to change the standardized of the human tissue regarding to the frequency we are dealing with.

The DUT which has a long arm, it is located under the hollow tub and close to the head shape due to our need to place the mobile phone in the edge of the head as it must be in the reality.

Finally, the measurement shouldn't be in one side of the human head and we should consider that the measurement of the SAR very sensitive so we can get a very big different result because of the complex nature of the near field.

4.2 The Techniques Of Reducing SAR

Reducing the electrical magnetic field of the antenna will effect to reduce the value of SAR in human head that can be made by using the RF shield to the front side of the phone. The RF shield which has been made of ferromagnetic material so it can help to reducing the value of the SAR because of the suppression of the surface current on the front of the mobile phone box as shown in figure 4.2.1.

During the travelling of the electrical magnetic wave through the free space the wave will be divided to three parts, the first part of the wave going to transmitted, other parts will reflect and absorbed, the function of the electrical magnetic wave material absorbed the energy of the wave as magnetic loss then convert it to energy which it can be feel as heat.

The main materials that are used for making electromagnetic shield are sheet metal, metal foam and material, the holes in the shield should be smaller than the wave length to keep it away.

The value of the reduction depends mainly on the material that have been used, the hickness of the shield, the frequency of the field, the shape, the directivity of the aperture to indicate the electromagnetic field and the size of the shield.

http://en.wikipedia.org/wiki/Electromagnetic_shielding

Figure 4.2.1 The RF shield in mobile phone.

The identification of the RF shield can be determined by determining the value of the SAR reducing factor (SRF) so if the value of SRF is very large that's mean its need to greater RF shield.

SRF can be calculating by the following formula:

$$\boldsymbol{SRF_{Total}}(\%) = \frac{\boldsymbol{P_{abs} - P_s}}{\boldsymbol{P_{abs}}} \times \mathbf{100} \qquad \textbf{(4.2.1)}$$

$$\boldsymbol{SRF_{1g}}(\%) = \frac{\boldsymbol{SAR_{1g} - SAR_{1g,s}}}{\boldsymbol{SAR_{1g}}} \times \mathbf{100} \qquad \textbf{(4.2.2)}$$

$$\boldsymbol{SRF_{10g}}(\%) = \frac{\boldsymbol{SAR_{10g} - SAR_{10g,s}}}{\boldsymbol{SAR_{10g}}} \times \mathbf{100} \qquad \textbf{(4.2.3)}$$

- (SRF) is the total SRF
- (P_{abs}) is the power absorbed in the head model, (P_s) is the power in the RF shield.

- (SRF_{1g}, SRF_{10g}) is the SRF for 1g and 10g peak SAR
- (SAR_{1g}, SAR_{10g}) is 1g and 10g peak SAR (without RF shielding)
- ($SAR_{1g,s}, SAR_{10g,s}$) is 1g and 10g peak SAR (with RF shielding)

Headset vs SAR

Based on the study of UK consumers association reported that the headset can reduce the value of SAR, because of the wire that's connect the headset to mobile phone therefore the wire is working as channeled unshielded radiation and antenna to the human head. The SARtest discovered that can reduce the absorbed radiation until 70%.

4.3 The Limit of SAR in Different Countries

TABLE 4.3.1 The limit of the SAR in different countries [13].

Country/Region	SAR measurement protocol	Reference to SAR limit	Limit
Europe	European Specification ES 59005 (1998)	ICNIRP Guidelines 1998 (ICNIRP 1998)	2.0 W / kg in 10g of tissue
Australia	Australian Communication Authority (ACA) standard (ACA RS 1999)	Australian standard AS/NZS 2772.1	1.6 W / kg in 1g of tissue
US	Federal Communications commission(FCC) Guidelines(FCC1987)	American Standard ANSI C95.1 (ANSI 1992)	1.6 W / kg in 1g of tissue

Regarding to the table 4.3.1 the average limit of European countries is 2.0 W / kg in 10g of tissue is absorbing the most signal, Australia and US has the same limitation 1.6 W / kg in 1g of tissue.

In India they were following the same limitation as Europe until 2012 then they have switched to follow the same limitation as US, and the test of India in SAR is done by Telecommunication engineering center. There are many countries have been following the same limitation of US such as Korea, Bolivia and Canada 1.6 W / kg in 1g of tissue.

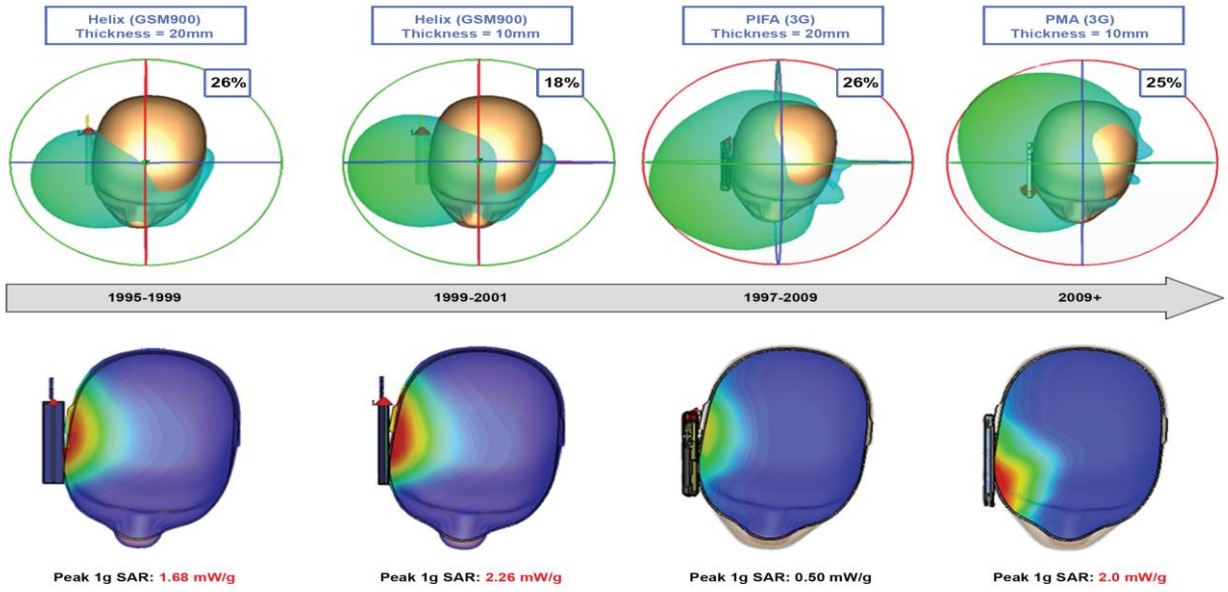
4.4 SAR With Different Types Of Antenna

There are some types of antenna that have a really good deal with SAR, the table 4.1.3 shows the dealing between the level of SAR with different types of antenna.

Table 4.4.1 The level of SAR in different types of antenna.

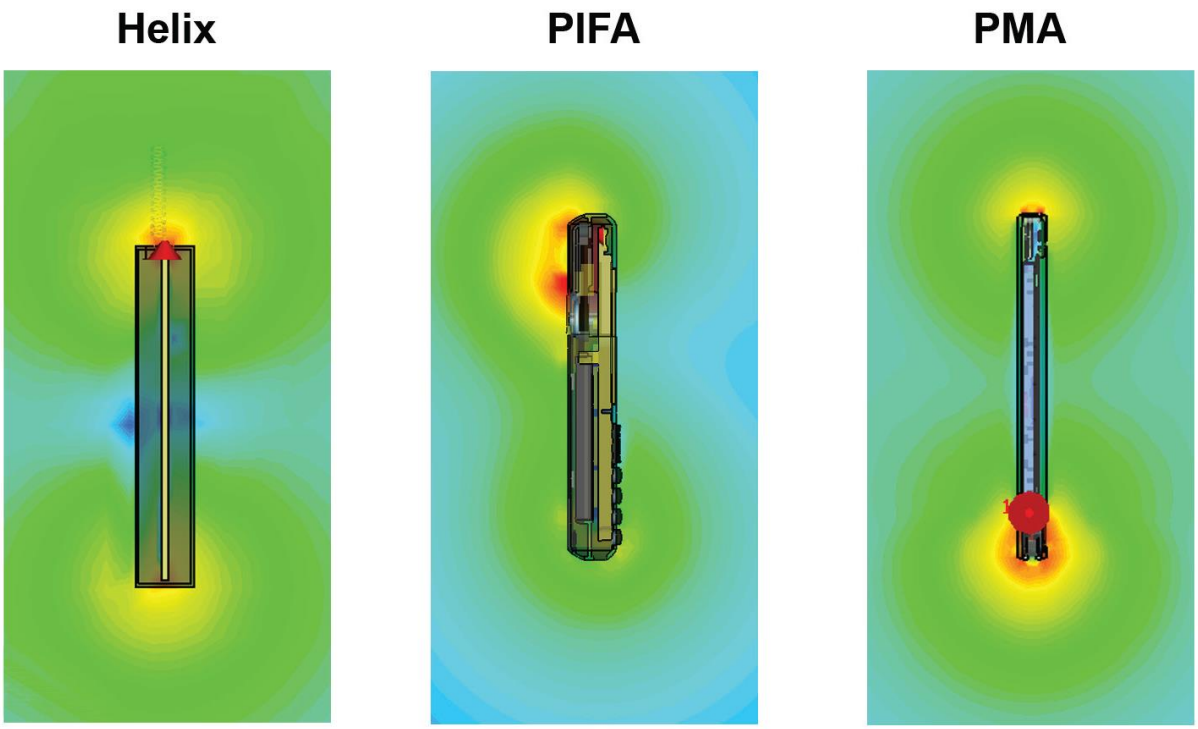
Type of antenna	Level of SAR
Helix	High SAR
Slot	LOW SAR
PIFA	LOW SAR
PMA	High SAR

As showing in figure 4.4.1 first shape show the human head with Helix external antenna which used since 1995 to 1999 with thickness 20 mm and the value of SAR 1.68 mW/g , we notice in the second shape it had used the same type of antenna but the thickness have changed from 20mm to 10mm because of this change of the thickness the value has increased to 2.26 mW/g, the used of PIFA from 1997 to 2009 which has a really good deal with SAR as we mentioned in table 4.4.1 and PMA which used above 2009 with SAR value 2.0 mW/g.



http://www.scribd.com/doc/254155865/Mobile-Antenna-Design#scribd

Figure 4.4.1 The evaluation and the performance of GSM antenna types from 1995 to 2011.



http://www.scribd.com/doc/254155865/Mobile-Antenna-Design#scribd

Figure 4.4.2 Comparing between the radiation of Helix, PIFA and PMA.

We can notice from the previous figure 4.4.2 that he radiation of Helix and PMA are like a dipole antenna but PIFA has different characteristics of radiation which guide us to notice that it is a type of microstrip antenna. So we can mention that the dipole antenna hasn't a good relation with SAR.

Table 4.4.1 The highest SAR values in different well known mobile phone manufacture companies [12].

Manufacture company	Model	SAR in W / kg
Apple	IPhone 5	1.25
Nokia	Lumia 630	1.52
Sony Ericsson	Z1010	1.41
Samsung	P400	1.18
Blackberry	Curve 9320	1.56

Table 4.4.2 The lowest SAR values in different well known mobile phone manufacture companies [12].

Manufacture company	Model	SAR in W / kg
Apple	IPhone 4,5c,4s,6	1.18
Nokia	9300	0.07
Sony Ericsson	Tz600	0.16
Samsung	X830	1.18
Blackberry	Curve 8900	1.01

Chapter 5

Problems of Mobile Phone Antennas Design

Mobile phones antenna design is a complicated process. There are a lot critical considerations, the place where the antennas will be installed in mobile phone chassis for example, and regulatory requirements, what are the possible bands of frequencies could be used. In addition, nowadays, mobile phones have multiple antenna for different functions.

5.1 The Radiation Efficiency of the Mobile Phone Antennas

The performance of an antenna depends on the antenna element itself and the ground plane of the mobile phone. That leads to the small size antenna problem, which means that there are limitations for how much the antenna can be minimized, in term of available space in mobile phone or in the expectable good performance. Mobile phone antennas have low efficiencies, and low input resistance and high input reactance, that's create a difficult for matching impedance between the antenna element and the transmission line.

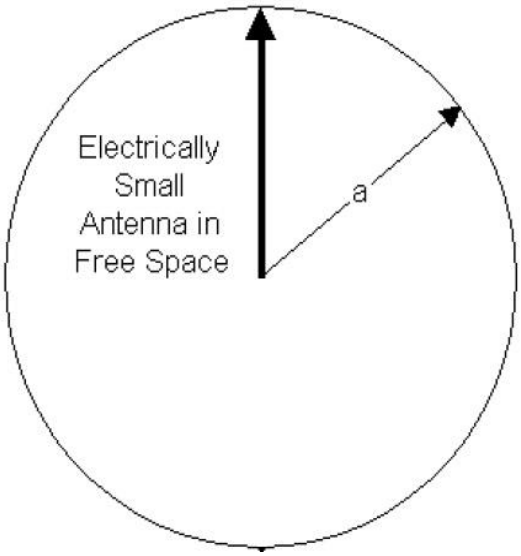


Figure 5.1.1 Imaginary Sphere encloses an antenna in free space with radius a.

At the beginning Wheeler introduced a method to define the maximum volume radiation for electrically small antenna by two measurements, first one in free space and the other within an imaginary closed sphere with radius $a = \frac{\lambda}{2\pi}$ as shown in figure 5.1.1 [18]. Wheeler considers that the antenna resonance frequency can be described as two series resistances; radiation resistance R_{rad} and loss resistance R_L (of used material) [18]. The radiation efficiency of a small antenna can be determined by the following formula:

$$\boldsymbol{\eta_{rad} = \frac{R_{rad}}{R_L + R_{rad}}} \qquad \textbf{(5.1.1)}$$

The current distribution in the small electrically antennas has small space to flow as it's obvious, that's make the antenna to behave as a capacitor, and consequently the current will be zero, that means there is no radiation power. In that case we must satisfy a proper matching impedance network in term to get radiation power by the small antenna. Radiation resistance for small antennas becomes very small due to weak radiation [26].

Efficiency of a system- which consists of antenna and matching network- can be expressed with help of the radiation efficiency which described in equation 5.1.2 by:

$$\boldsymbol{\eta_s = \eta_{rad}\eta_m} \qquad \textbf{(5.1.2)}$$

η_s: Efficiency of the system.

η_m: Efficiency of the matching network.

Based on the equations 5.1.1 and 5.1.2 we can conclude that in order to get higher efficiency of a system, we have to take in account several parameters:

1. Maximum Radiation resistance.
2. The lowest value of loss resistance.
3. The lower mismatch network as possible (lower |Γ|).

In addition, the mobile phone case's material type, and the surrounding environment (such as hand-held, which it will be discussed later) have an additional effect on the radiation efficiency. Antennas with bigger size have better efficiencies than small antennas. On other hand the space which is available for antenna in chassis of mobile phones is very restricted. Thus there is always tradeoff between size and performance.

5.2 Bandwidth

We can determine the bandwidth of an antenna by determining the impedance for all frequencies in that desired range. But in small antennas case the bandwidth depends on others factors. Small antennas such as PIFA have less real input resistance (it approaches to zero) and highly reactive input impedance, which leads to mismatching network problem. To match the impedance of a small antenna with its feeder's impedance is a quite critical issue. The bandwidth depends on the reflection coefficient and matching networks.

The quality factor Q is a limit of the accessible impedance bandwidth at certain efficiency. Sometimes the small antenna is defined by Q.

As it's mentioned in the previous chapters, we can define the bandwidth by the following formula:

$$\boldsymbol{Bw} = \frac{\boldsymbol{f_{max} - f_{min}}}{\boldsymbol{f_0}} = \frac{\mathbf{1}}{\boldsymbol{Q}} \qquad \textit{Valid for } Q >> 1 \qquad \textbf{(5.2.1)}$$

For obtaining wide bandwidth we need to have the most possible minimum value of Q that's a hard challenge to solve. In 1947 Wheeler started to define radiation power factor, and the work on that issue is continued until nowadays. The most famous paper was published by Mclean 1996 [90] which describe the Q factor for a linearly polarized in free space based on Wheeler's concept as following:

$$\boldsymbol{Q_{LP} = \frac{1}{k^3 a^3} + \frac{1}{ka}} \qquad \textbf{(5.2.2)}$$

Where, $k = \frac{2\pi}{\lambda}$

For circularly polarized antenna Q has slightly different requirement:

$$\boldsymbol{Q_{CP} = \frac{1}{2}\left[\frac{1}{k^3 a^3} + \frac{1}{ka}\right]} \qquad \textbf{(5.2.3)}$$

The small antennas with higher values of Q have narrow bandwidth and that due to the low radiation resistance and high reactance. The Q factor for a small antenna is defined as following:

$$\boldsymbol{Q_{minimum} = \frac{2\omega max(W_e, W_m)}{P_{rad}}} \qquad \textbf{(5.2.4)}$$

ω: The angular frequency.

W_e: The stored electric energy.

W_m: The stored magnetic energy.

P_{rad}: Radiation power.

With assuming that's there is no stored energy (W_e, W_m) inside the sphere (figure 5.1.1) which means reactive fielded equal to zero.

Based on equation 5.2.4 we note that's, mathematically to get the lowest value of Q the power radiation should be maximized, that's lead us to the efficiency problem. The mobile phones antenna design's problems are related with each other. Consequently more challenges are added to design process.

Multiband antenna can be used in order to solve the bandwidth limitations problem partially, since there is no small multiband antennas can cover all desired bands. As obvious from the discussion above, the wide bandwidth is a hard challenge to achieve by an ESA due to low profile (small size). In modern mobile phones, there are several antennas for different functions as mentioned in mobile phones antennas design's chapter. Each antenna operates with certain band; for example one antenna for higher band frequency, and one for GPS etc. Then all desired bands can be covered by that method. On other hand we have to take care on matching network, since we have several antennas to deal with, mutual coupling antenna to antenna loss, and the available space in mobile phone's chassis.

5.3 Mutual Coupling Antenna to Antenna Loss

The mutual coupling is the amount of the absorbed energy by an antenna when another nearby antenna is operating [4]. This interaction between near antennas is unwanted because that's amount of the absorbed energy should be radiated instead. Thus the mutual coupling reduces the efficiency of the antenna on both receiving and transmitting mode.

Mobile phones consist of several antennas, and with mobile phones geometrical dimensions, we find that the antennas near to each other in such way make the mutual coupling loss inevitably exists. So a good isolation is required to keep the antennas in mobile phone's chassis far from each other as possible. For that the antennas are generally distributed on top and bottom of the chassis which reduce the coupling loss. The coupling loss can be 1-2dB for antenna efficiency. Isolation values for smartphones which have same ground plane at the low band are about 10 dB, and can be 20 dB for the high band [28]. The coupling loss is not constant, it's varying; the coupling loss at low frequencies and at high frequencies are different.

The isolation can be increased by:

- Minimizing the correlation coefficient between the antennas.
- Increasing the separation space between antennas as much as possible.
- Using different polarizations for the antennas.

- Using proper filters with the antennas to decrease the effects of opposite antennas frequencies.
- Using dielectric walls between the antennas.

The isolation between two antennas is measured by connecting both of them to a Vector Network Analyzer and measuring of S12.

5.4 The Hand-Held Environment Problem

Table 5.4.1 Dipole performance at different locations of simulated distance from human head [24].

Distance from Head	**cm**	**Free space**	**5**	**4**	**3**	**2**	**1**
Input Impedance	Ω	75+j1.3	64+j27	60+j28	62+j35	52+j17	60+j23
Radiation Efficiency	**%**	100	72.3	63.4	50.5	42.7	29.1
Max. Directive Gain	**dB**	2.15	4	2.15	0	-1	-2
Min. Directive Gain	**dB**	2.15	-7	-7.5	-8	-9	-12

As shown in table 5.4.1, when the antenna is closer to the human head, the input impedance decreases, consequently the resonant frequency of the antenna decreases also. The absorption which occurs by human head reduces the radiation efficiency to 29% when the distance 1cm as shown above.

Other effect of handheld is in the directivity of the antenna. The difference between two positions at free space and 5cm far from the head, shows that the directivity increases from 2.15 dB to 4 dB, that's because the sum of radiation fields and fields scattered by the head which somehow compensate the loss in radiation of efficiency[24].

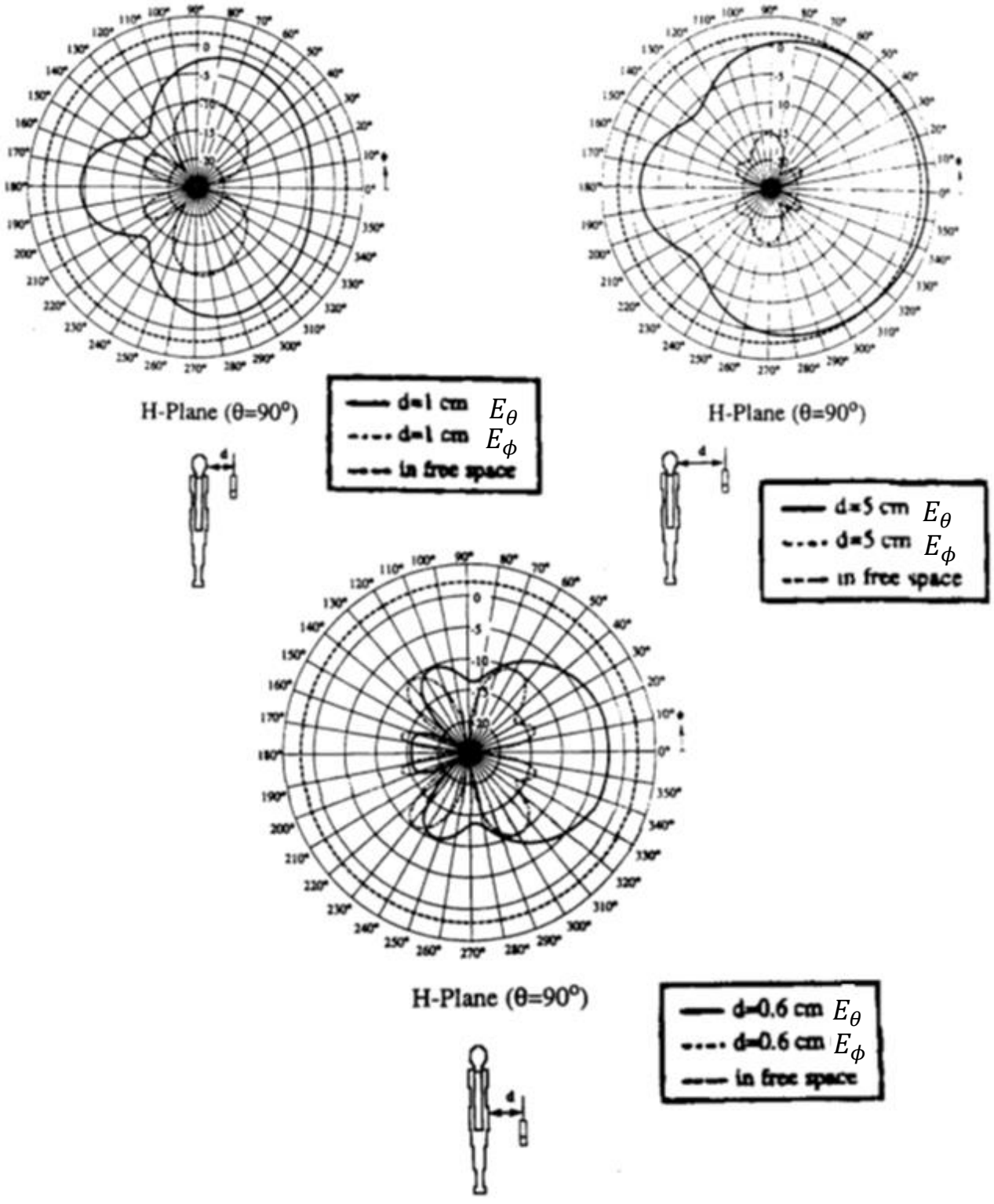
In general we can conclude that when the mobile phone is closer to the head the gain will fall quickly and the directivity is lost due to polarized fields.

Another model of a dipole antenna with a simulated distance is 0.6 cm from human head, and by applying the method of moments (***MoM***). The input impedance is 65 + j13 Ω, directional properties are same as 1cm and 15% radiation efficiency [25].

In figure 5.4.1 we can note that all three of the patterns radiation show -10dB nulls directly behind the user because absorption of energy by the modeled tissue, the attenuation increases as distance from the body decreases, and is worst for operation near the human torso [24].

Table 5.4.2 The effects of iPhone4 cases on antenna efficiency and absorption in the user's hand [17].

Case Type	Efficiency		Hand Peak SAR (mW/g)	
	GSM900	3G	GSM900	3G
No Case	14%	16%	19.3	27.0
Plastic Case (t = 1 mm)	16%	17%	11.6	12.6
Plastic Case (t = 2 mm)	18%	17%	7.1	8.0
Plastic Case (t = 4 mm)	18%	17%	2.3	5.4
Ring: Plastic (t = 1 mm)	16%	16%	12.7	11.3
Ring: Plastic (t = 2 mm)	18%	16%	7.9	8.5
Metal Case (t = 1 mm)	0%	1%	0.0	0.2
Ring Metal Case (t = 2 mm)	0%	3%	0.0	0.0
Chrome Plated Plastic (t = 2 mm)	4%	7%	9.2	11.7
Ring: Metal (0.3 mm) + Air (3 mm)	23%	15%	16.3	25.8
Ring: Metal (0.3 mm) + Air (1 mm)	11%	17%	3.6	21.5
Ring: Plastic (1 mm) + Air (3 mm)	24%	22%	7.4	9.8
Ring: Plastic (1 mm) + Air (1 mm)	22%	21%	7.9	9.6



Chuang, H. R. "Human Operator Coupling Effects on Radiation Characteristics of a Portable Communication Dipole Antenna"., IEEE Transactions on Antennas and Propagation, v. 42, n.4, April 1994, pp. 556-560.

Figure 5.4.1 .Free space dipole radiation pattern compared to the calculated (using MoM) radiation patterns for operation of a dipole antenna near the simulated human head and torso.

Coupling of energy to the hand of the user is another problem caused by hand-held environment. Figure 5.4.1 shows the experimental and measured results of the reflected power radio at different distance from the human hand for internal PIFA antenna by using an FDTD model for human hand. As seen below in figure 5.4.2 when the distance d becomes less more and more (the hand of the user gets closer to the antenna) the resonance frequency is shifted, and the reflected power radio becomes more, which means that's the antenna's efficiency is obviously degraded.

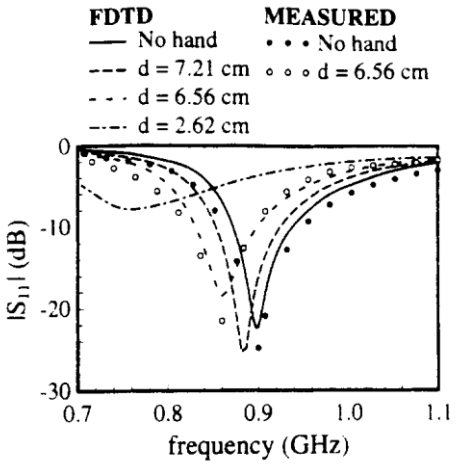


Photo was loaded from: Jensen, M., Rahmat-Samii, Y. "EM Interaction of Handset Antennas and a Human in Personal Communications".

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Figure 5.4.2 The experimental and measured results of the reflected power radio at different distance from the human hand.

The hand-held environment problems add more challenges in design process. Since it's so important to take in account the different scenarios. Some of that's issues can be solved by design and installed the antenna in the lower part of the mobile phone to keep the antenna far as possible from the hand of user. Another critical factor can play big role in this term is a good isolation for antenna.

IPhone 4 suffers from isolation antenna problem. Many users reported that signal strength of the phone was reduced when touching the lower left of the phone, which causes dropped calls in some areas [26]. On other hand Apple recommended the consumers to do not grip the phone at that corner.

On July 16, 2010, Steve Jobs advertised that Apple would offer a case for the consumers to help solve the antenna issue [26]. Table 5.4.2 shows an iPhone4 and the effects of different material of casing on efficiency and SAR. Metal cases with thickness equal to 1mm, and 2mm, keep the efficiency without any effects and SAR values at almost zero.

Summary and Conclusion

Mobile phone antennas have several critical parameters such as the geometric dimensions, dialectical materials which are used in design, and the ground plane of the antenna. All of that should be taken into account for optimizations process. The fast development of mobile phones devices add more demand on optimizations of the antennas.

In this thesis we investigate PIFA antenna type by design two models of it. The first model handles one single band with resonant frequency 1.575GHz which could be suitable at that frequency for GPS signal with reasonable gain from -6dBi to 1.5dBi and with return loss (S11) -13dB. The bandwidth 1% which is sufficient since there is no bandwidth practically.

The second proposed PIFA is a wideband antenna with dimensions $24 \times 10 \times 4mm^3$cover range from 1800MHz to 2600MHz. This important range covers GSM (1800MHz & 1900MHz), UMTS (2100MHz), Bluetooth & Wi-Fi (2.4GHz), and LTE system (2.3GHz, 2.5GHz, and 2.6GHz). The Bandwidth is 36.36% which is considered as a wide bandwidth. This wideband PIFA antenna is designed using slotted ground plane and air as a dielectric material between the planar element and the ground plane.

Mobile phone antennas design is a big challenge. Mobile phone antennas designs have lot problems. Low efficiencies can be the most critical problem which leads to tradeoff between size and performance.

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High-Performance Indoor VHF-UHF Antennas:

Technology Update Report

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1.0 Introduction and Summary of Findings

In 1995 MegaWave Corporation, under an NAB sponsored project, developed a broadband VHF/UHF set-top antenna using the continuously resistively loaded printed thin-film bow-tie shown in Figure 1-1. It featured a low VSWR (< 3:1) and a constant dipole-like azimuthal pattern across both the VHF and UHF television bands.

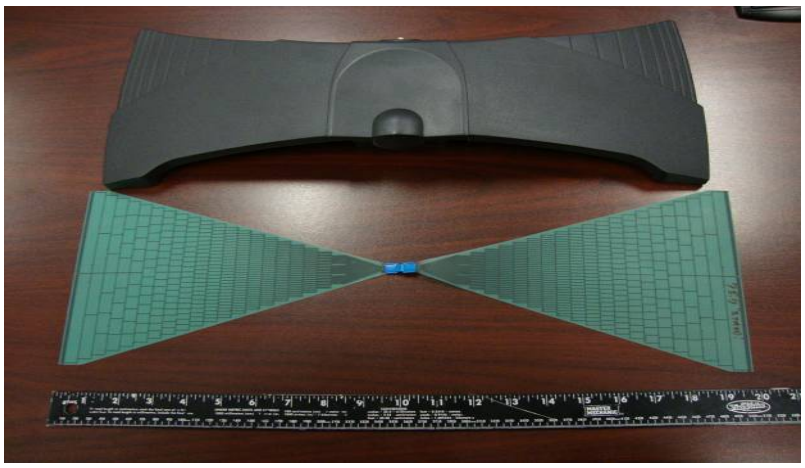

Figure 1-1: MegaWave 54-806 MHz Set Top TV Antenna, 1995

In the 15 years since then much technical progress has been made in the area of broadband and low-profile antenna design methods and actual designs. These improvements have been published in: technical textbooks, peer-reviewed articles, patents, government research and development reports, and seminar proceedings. As a developer of advanced antenna systems, primarily for the U.S. government, MegaWave constantly reviews these sources and acquires the latest computer based EM simulation tools in order preserve its competitive advantage. In this project, this knowledge was used to identify ten candidate design methods and technologies that have the potential to materially improve the performance of indoor VHF-UHF TV antennas. This report describes each candidate and its potential to improve indoor ”set-top” reception of DTV signals between 54 and 698 MHz.

Of course, it must be kept in mind that, while advanced design methods and actual physical designs exist, so do the laws of electromagnetics. Maxwell’s equations have resulted both in practical as well as, what Dr. R. C. Hansen humorously calls, “Pathological Antennas”. These pathological designs are described in his most recent textbook [1], especially in the area of electrically-small and broadband designs. It is instructive to apply these fundamental limitations to the problem at hand, the set-top TV antenna.

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Consider that a half-wavelength in the low VHF TV band varies between 9.2 and 5.6 feet; between 34 and 27 inches in the high VHF band and between 12.6 and 8.5 inches in the UHF (470-698 MHz) band. A dipole antenna whose physical length is less than its wavelength divided by pi (λ/π) is considered to be an electrically "small" antenna (ESA). ESAs unfortunately are characterized by narrow bandwidths and low gains. Assuming 2 to 3 feet as a maximum acceptable length for an indoor or set-top antenna, it definitely falls into the ESA category in the low VHF band. But, in addition to size constraints and the resulting difficulty in obtaining acceptable performance from a single antenna over the 54 to 698 MHz spectrum, there are other concerns. Indoor and set-top antennas are fundamentally disadvantaged due to building penetration losses and by proximity to sources of manmade radio noise. The former effect is more pronounced at UHF and the latter at low VHF channels. Both can have a significant deleterious effect on antenna performance. This brief discussion highlights the difficult problems inherent in designing efficient, high performance antennas for the indoor/set-op TV environment. Fortunately, emerging technologies may effectively address these concerns.

This report is organized as follows. Sections 2.1 through 2.10 contain summaries of each advanced method and hardware technology identified as a potential candidate for high-performance indoor VHF-UHF DTV antennas. Each section includes a list of references and, in many cases, photographs and performance data for multiple implementations of the technology that is described. Section 3 includes conclusions and a conceptual design for a practical indoor/set top VHF-UHF antenna system.

The authors evaluated each technology and arrived at the conclusions and design concept after sorting the nine hardware candidates into three categories as follows:

- Mature technologies that **do not** require any CE-909-A channel designator or signal quality information from the DTV receiver:
  - Fragmented Antennas (Section 2.2)
  - Non-Foster Impedance Matching (Section 2.3)
- Mature technologies that **do** require channel and quality data from the receiver:
  - Active RF Noise Cancelling (Section 2.4)
  - Automatic Antenna Matching Systems (Section 2.5)
  - Physically Reconfigurable Antenna Elements (Section 2.6)
- Emerging technologies that show promise, but are not sufficiently mature or practical at this time:
  - Metamaterials (Section 2.7)
  - Electromagnetic Band Gap (EBG) Materials (Section 2.8)
  - Fractal/Self Similar Antennas (Section 2.9)
  - Retrodirective Arrays (Section 2.10)

A common thread connects each of these technology areas: ***advanced computational methods***. Whether a particular technology is mature and immediately applicable or emerging and highly speculative, various schemes for antenna design optimization are universally applicable and described in Section 2.1. These methodologies apply to all of the candidate technologies discussed in Sections 2.2 through 2.10, and accordingly was placed at the beginning of Section 2. If even one of the optimization algorithms described had been available during the development of MegaWave's 1995 broadband set top antenna, it is likely that markedly better gain performance would have resulted, especially in the low and high VHF bands. Another attractive and potentially very significant capability offered by optimization algorithms is the possibility of discovering entirely new antenna geometries, rather than simply optimizing a pre-existing geometry.

Table 1-1 ***subjectively*** ranks the nine identified candidate hardware technologies (2.2 - 2.10). A score of 10 represents perfection. By maturity we mean how close to off-the-shelf a particular technology's hardware is and how well it basic principle of operation has been vetted in the literature. The term SWAP refers to size/weight and power.

	Method/ Technology	Active/ Passive	Requires CE-909-A Interface	Maturity	Vetted	Risk	Design Complexity	SWAP	Comments
2.1	Adv. Comp. Methods	N/A	N/A	9	9	Very Low	N/A	N/A	Applies to all Technologies
2.2	Fragmented	Passive	No	7	7	Low	Low	9	Planar
2.3	Non-Foster	Active	No	6	7	Low	Moderate	7	Limited Bandwidth
2.4	Active Noise Cancelling	Active	Yes	7	6	Low	High	3	Requires I&Q
2.5	Automated Antenna Matching	Active	Yes	7	7	Moderate	High	7	Requires complex TV interface
2.6	Reconfigurable Antennas	Active	Yes	6	7	Moderate	High	6	Control of MEMS w/DC
2.7	Metamaterials	Passive	No	3	4	High	High	8	Emerging/ Availability an issue
2.8	EBG	Passive	No	5	6	High	High	5	Inherently Narrow Band Maybe useful for shielding
2.9	Fractal/Self Similar	Passive	No	6	4	Moderate	Moderate	8	Controversial Performance Gain
2.10	Retrodirective	Active	Yes	4	5	Very High	Very High	2	Narrow - Band, Large

Table 1-1: Candidate Technologies Considered and Their Ranking

As an example of how advanced computational methods could be combined with an advanced hardware technique, that does not require a CE-909-A interface, is described at the end of Section 3 and summarized here.

Using the genetic algorithm described in Section 2.1.5 a fragmented antenna was designed and combined with a non-Foster-matching circuit to provide a planar 54-698 MHz dipole approximately 13 by 13 inches with significantly better gain, especially in the 54-88 and 174-216 MHz bands, than the 1995 MegaWave/NAB set top antenna. Figure 1 shows the broadband fragmented planar element's design obtained after approximately 24 hours of computational time on a PC. Details of the specific method used are in Section 2.2 of this report. It is well matched across the UHF DTV band, but requires some passive matching in the high VHF band (which would also serve as the band combiner) and the more robust matching capability of the active Non-Foster-Matching technique, described in Section 2.3, for the low-VHF band.

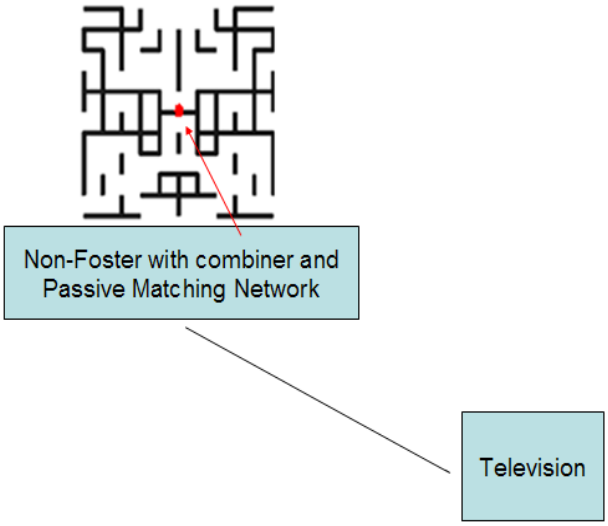

Figure 1. 13 x 13 Inch Planar Fragmented Non-Foster Matched VHF-UHF Antenna

An omni-directional version could also be designed. It should be stressed that the above is included here only to illustrate the notion of combining advanced computational broadband antenna element designs with emerging electronic antenna matching capabilities and that other antenna element geometries are also possible, depending on the starting conditions, trade space dimensions and performance goals provided to the optimizer.

The authors want to make clear that 90 percent of the techniques and ideas contained in this study are the work of others, as published in the open literature and referenced herein.

2.0 Specific Design Methods and Hardware Technologies Investigated

2.1 Advanced Computational Methods

2.1.1 Summary

Optimization methodologies abound, and they are extensively used in every aspect of engineering design, in particular antenna design. Optimization algorithms are useful in two ways. They can be used to optimize the design parameters for a user-specified antenna geometry (for example, element spacing, length and diameter in a Yagi-Uda array). They also can *generate designs* that are impossible to achieve otherwise. In both cases, optimization involves meeting specific performance objectives (typically, VSWR, gain, bandwidth, and so on).

Optimization algorithms have become progressively more important as the limitations of classic analytical techniques have become progressively more apparent. While the equations underlying electromagnetic theory are well understood and accurately describe all electromagnetic phenomena, in most practical cases they cannot be solved analytically or, oftentimes, even numerically. Designing better antennas requires improved methodologies, and state-of-the-art optimization algorithms have proven very effective. There is no question that these techniques are applicable to the set-top antenna design problem, and that they should receive considerable attention in future design activities.

There are many different optimization methodologies that fall into two broad categories: analytical methods and heuristic methods. Analytical methods are based on precise mathematical formulations of the optimization problem. Even though they may be fundamentally numerical in nature, they involve standard mathematical operations such as computing derivatives or evaluating integrals. Heuristic methods may involve equations, but the equations are not the result of an analysis. Instead, they are offered without "proof" based on the fact that they "work."

Many optimization heuristics are Nature inspired. The steps an algorithm performs to optimize an *antenna* are based, for example, on how *bacteria* forage for food. As disparate as these entities may seem, there is a connection, at least in the sense that bacteria finding a good food source is similar to finding an antenna with a good gain-bandwidth product. Optimization algorithms of this type are usually referred to as "metaheuristics," a term intended to emphasize that the method is both *empirical* and *conceptual* in nature. Thus, an effective bacteria foraging algorithm can be implemented in many different ways because the bacteria foraging metaheuristic simply suggests an analogy to Nature that is implemented in a computer algorithm working on

an antenna problem. The metaheuristic thus is an *algorithmic framework* instead of a list of steps or instructions.

Several Nature inspired metaheuristics are described. A brief summary of each algorithm is provided, and several example antenna problems solved by a variety of algorithms are discussed. The algorithms include Ant Colony Optimization (ACO), Particle Swarm Optimization (PSO), Genetic Algorithm (GA), Simulated Annealing (SA), Central Force Optimization (CFO), Invasive Weed Optimization (IWO), Intelligent Water Drop (IWD) algorithm, and Bacteria Foraging Optimization (BFO). There are many other optimization algorithms [for example, Space Gravitation Optimization (SGO), Integrated Radiation Optimization (IRO)], but they have not been applied to antennas or antenna related problems.

Each of these algorithms, except one, is inherently *stochastic* because its Nature inspired algorithmic model relies on randomness in its functioning. The underlying equations contain true *random variables* whose values are computed from a probability distribution and consequently cannot be known in advance. As a result, every time a stochastic optimizer run is made, its results are different than the previous run even when exactly the same run setup parameters are used. The performance of stochastic optimizers is necessarily characterized statistically (for example, average values, standard deviations). This may be a limitation in the utility of optimization algorithms if they are used in a set-top antenna on a real time basis. For example, a self-structuring antenna (SSA) must reconfigure itself in real time in response, for example, to a changing environment.

The one algorithm that is not inherently stochastic is Central Force Optimization (CFO) whose Nature inspiring metaphor is gravitational kinematics, the branch of physics that deals with the motion of masses moving under the influence of gravity. The underlying equations are Newton's equations of motion, which are completely *deterministic*. CFO analogizes these equations in "CFO space" by flying "probes" that are similar to small satellites to search a decision space "landscape" for the maximum (optimal) values of a function (for example, antenna gain as a function of element length and polar angle). CFO has been applied to antenna design and network synthesis, and tested against many recognized benchmark functions used to evaluate optimization algorithms. It therefore may be especially useful for the set-top antenna problem.

2.1.2 Introduction

This section describes developments in antenna design optimization over the past fifteen years or so that have been driven largely by the availability of progressively more powerful computers. A plethora of new optimization algorithms have been

introduced and tested and are now in widespread use. The new antenna designs often are non-intuitive, occasionally even counter-intuitive, but all share the common feature of not being accessible in any other way. State-of-the-art optimization algorithms can effectively solve intractable problems that have no analytical solutions or are too complex to apply traditional analytical techniques. These approaches are useful right now in designing set-top television antennas, and they will continue to be useful whatever form future set-stop systems take. Some of the more important and interesting optimization algorithms are described here.

Optimization Methodologies. The problem of locating the maximum values of a function is generally referred to as "multidimensional search and optimization." As pointed out above, any problem involving three or more design parameters ("decision variables") is a multidimensional problem, and simple methods such as plotting the function to be maximized cannot be used. Methods for solving these problems fall into two broad categories: analytical methods and heuristic methods. Analytical methods, which involve computing derivatives and gradients, are of limited use, especially in the complex landscapes associated with antenna design. Stringent performance requirements in terms of bandwidth, radiation pattern, and standing wave ratio (SWR) make antenna optimization problems particularly difficult because the landscape is usually extremely multimodal with narrow resonances and often high sensitivity to slight parameter variations. Heuristic optimization methodologies, which are inherently numerical in nature, are effective in dealing with these issues, and consequently they are considered here while analytical approaches are not.

An entire class of heuristic optimization algorithms are "Nature inspired", and these appear to be the most effective. A Nature inspired algorithm is a computer search and optimization program whose function mimics some natural process. These programs are described as being "metaphorical" because they analogize some natural process without precisely modeling it. For example, "Ant Colony Optimization" (ACO) is an algorithm that simulates (to some degree) the behavior of ants seeking food. Thus, ACO is *inspired* by the *metaphor* of ant foraging. All such algorithms *evolve* a solution to the optimization problem over a series of steps or iterations, and almost all such algorithms are *stochastic population*-based methodologies. An initial population (of ants, for example) randomly (stochastically) moves through the decision space (landscape) step-by-step (iterating) in such a way that it converges on the largest food supply (maximum function value). The ants' progress is controlled by a set of equations that mimic real ant behavior in Nature. There are many Nature inspired algorithms, ACO being one of the earliest ones. The more important algorithms are discussed below with examples of their application to antenna optimization.

2.1.3 Ant Colony Optimization

Figure 1 illustrates the basic idea behind Ant Colony Optimization (ACO) [1]. The irregular objects represent the ants' nest (bottom) and a desirable source of food (top). It has been observed that ants seeking food eventually traverse the shortest path between the nest and food by marking that trail with a chemical pheromone that each ant can sense (probably by smell). If the path is unobstructed [(a) in the figure], then the ants simply walk a more-or-less straight line between home and the food supply. But, if an obstruction is imposed [(b) and (c) in the figure], then more ants eventually end up on the shorter trail between the food and the nest, which in turn results in a greater pheromone concentration along that "optimal" trail. By depositing progressively more pheromone on the shortest path, almost all of the ants eventually end up on that path, and the "best" solution has been found. The red lines in the bottom part of the diagram illustrate the path evolution with the eventual result that the shortest path is identified.

The ACO algorithm mimics the ants' behavior using equations that represent the *random* motion of individual ants subject to their pheromone environment. Instead of searching for food, the metaphorical ACO ants search the landscape of a decision space for the maximum value of the function to be maximized. But the process they follow is a simplified model of ant behavior as observed in Nature. And, just as real ants eventually discover the best food source, ACO's "ants" eventually converge on the function's global maximum value.

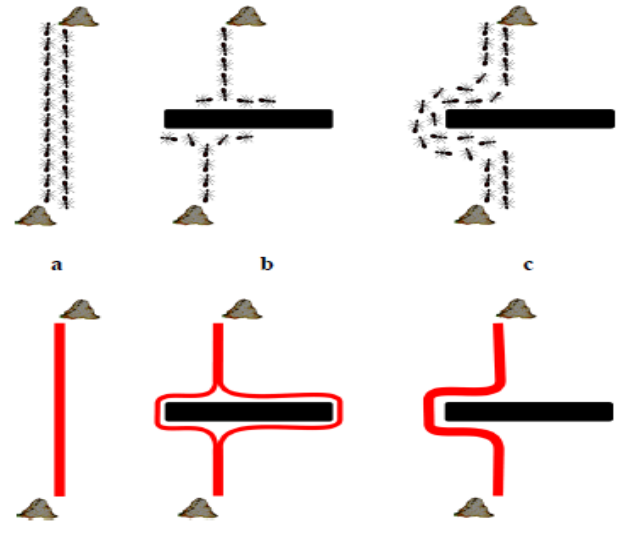

Fig. 1 Ant Colony Optimization: ants searching the shortest route using a pheromone trail.

Figure 1. Ant Colony Optimization Metaheuristic (reproduced from [1]).

2.1.4 Particle Swarm Optimization

Particle Swarm Optimization (PSO) [2] is another stochastic population-based Nature inspired evolutionary algorithm. PSO analogies the swarming behavior of fish or bees seeking food. Unlike ACO in which each "ant" creates a pheromone trail for other ants to follow, PSO's population of "agents" collectively communicate two pieces of information: each individual agent's "best" solution (greatest food concentration) and the population's overall (global) best solution. Equations that mimic bee and fish swarming then control each agent's subsequent motion in the decision space based on the competing tendencies of moving toward the global best and randomly exploring the vicinity of its best solution. As shown in Figure 2 for bees swarming around a flower concentration, after many steps PSO agents converge on the global best solution (highest flower concentration) because the local search fails to reveal any better solutions.

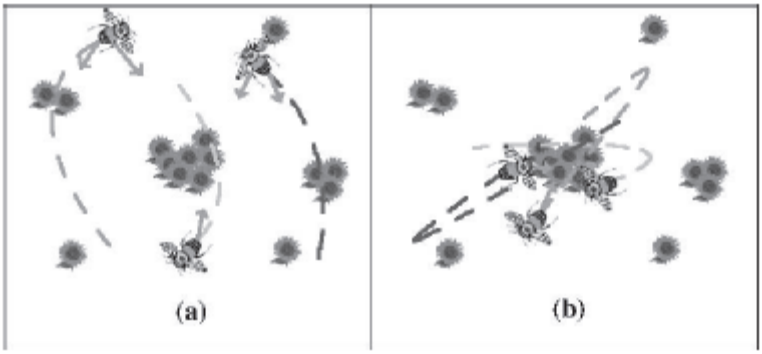

Figure 2. Particle Swarm Optimization metaheuristic (reproduced from [2]).

2.1.5 Genetic Algorithms

A Genetic Algorithm (GA) [3] analogizes the process of natural evolution or "survival of the fittest." When biological parents "mate," they exchange DNA to create a new individual ("child") whose characteristics are drawn from both parents by combining the parents' DNA. A GA creates successive generations of children who then serve as parents for the next generation whose children, in turn, will exhibit better "fitness" than the previous generation. In the context of search and optimization, the fitness is the value of the function to be maximized, so that the "best" fitness corresponds to the function's global maximum. As the GA progresses generation after generation, the best discovered fitness improves and eventually converges on the function's global maximum.

Figure 3 shows a typical GSA flowchart for an antenna optimization algorithm. It starts with a definition of the decision space (parameters to be optimized) and the "fitness function" to be maximized (for example, antenna directivity, or some specified combination of performance parameters such as gain, bandwidth, and so on). An initial population of "individuals" is randomly created, and each individual is defined by a *chromosome* that may be a binary sequence or a real number. Each chromosome comprises a set of *genes*, and each gene is one of the design parameters. For example, if the three design parameters were element length, inter-element spacing, and element diameter in a four element Yagi-Uda array, then there is a total of eleven design parameters, and each one is a gene. Thus, the optimization problem is defined on an 11-dimensional decision space, and the objective is to determine each of the eleven parameters so as to maximize some specific *fitness* function, say, the array's gain-bandwidth product. A separate computer program is used compute the fitness at each step for each chromosome (the "evaluate fitness" box in Figure 3).

After the initial population's fitnesses are evaluated, the "selection" process chooses two parent chromosomes that will mate ("crossover") to produce two children chromosomes in the next generation. The selection and crossover processes take many varied forms. For example, the selection of parents may be random, or "best mates worst," or best pairs pair wise through the population, and so on. The crossover operation likewise can take many forms. For example, the parents' chromosomes may be split at the midpoint with first and second parts being swapped, or a random break point might be used, or some other combinatorial approach taken. Finally, the children thus created are subject to some level of *mutation*, a random perturbation of the chromosome structure just as real chromosomes are mutated in Nature. The steps described thus far are essentially common to all Gas, but the next step in the flowchart ("elitist model") is not. In this GA, the worst individual in the new generation is replaced by the best individual from the previous generation, thus preserving the best solution from generation to generation as the algorithm progresses. As a final step, the best fitness is tested for convergence, and the process repeated until convergence is achieved.

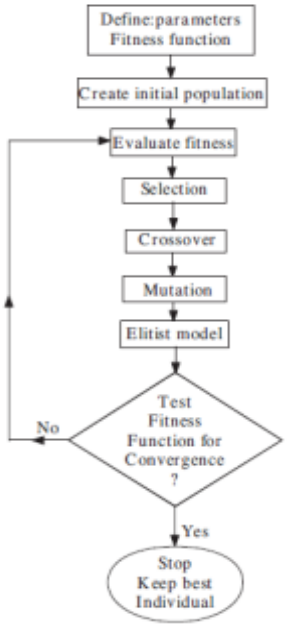

Figure 2. Flow chart diagram of real-coded GA.

Figure 3. GA flow chart (reproduced from [3]).

2.1.6 Simulated Annealing

Simulated Annealing (SA) [4] is a stochastic algorithm based on a metaphor drawn from physics instead of biology, as ACO, PSO, and GA are. SA analogizes the statistical mechanics of physical systems in thermal equilibrium with many degrees of freedom. In particular, the physical processes involved in annealing a solid as it cools forms the basis of the SA optimization algorithm, which has proven effective in optimizing problems with large numbers of decision variables. Because of SA's complexity, the algorithm is not described in detail. Instead its performance against a classic test problem is discussed.

The Traveling Salesman Problem (TSP) is a recognized example of combinatorial optimization that SA was used to solve because it constitutes a good test of an algorithm's effectiveness. The salesman must visit *N* different cities once each and return to his starting point. The problem is to determine the least costly route using a "cost" or "objective" function that is specified beforehand. Minimizing the cost is the same as maximizing its negative (note that minimization and maximization problems are exactly the same except for multiplying the objective function by -1). The TSP is a multidimensional search and optimization problem in the same vein as an antenna optimization problem, so that an algorithm suitable for one very likely is applicable to the other.

For the SA test, the TSP cost function is simply the total distance travelled by the salesman (to be minimized). Two different distance metric can be used, the standard Euclidean distance ("square root of the sum of the squares"), or the "Manhattan" metric (sum of the separations along the two coordinate axes), the latter being used in this case because it is simpler (less computationally intensive). Evolved solutions for TSP appear in Figure 6 and show a clear tendency towards removing redundancy in the travelled route, with the final solution (d) being close to optimal as discussed in [4].

Figure 4. Evolution of SA solutions to TSP (reproduced from [4]).

2.1.7 Central Force Optimization

Central Force Optimization (CFO) [5] is a new algorithm that departs significantly from all other Nature inspired metaheuristics. ACO, PSO, SA, and the other algorithms described below are all inherently *stochastic*. Every run with the same setup parameters in general produces a different set of solutions. No two runs yield the same results because these algorithms rely on true randomness in their functioning. The values of certain key variables in the algorithm are, by definition, random variables that are computed from a probability distribution. The values of these variables must vary from one calculation to the next, and their values are completely unknown and unknowable until the probabilistic calculation is performed.

CFO is quite different. It is based on an analogy drawn from gravitational kinematics, which in turn is based on Newton's laws of gravity and motion. Newton's laws are mathematically precise (completely *deterministic*) and, as a result, so too is CFO. CFO searches the decision space by "flying" probes through it whose trajectories are computed by deterministic equations that analogize Newton's laws of motion.

Figure 5 shows how CFO's probes move through a 3D decision space at each time step sampling the decision space by computing the fitness of the function to be maximized (shown by the darkened circles). CFO thus provides some major advantages over stochastic algorithms, *viz*, every run with the same setup returns exactly the same answers, and because of that characteristic only one run is necessary (stochastic algorithms usually are run many times and the results averaged). CFO has been effectively used for antenna optimization, and it holds considerable promise for use in set-top antenna design.

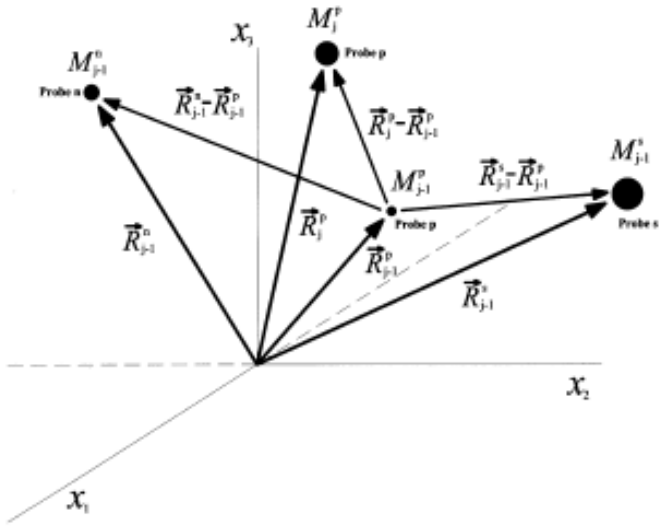

Figure 5. Central Force Optimization metaheuristic (reproduced from [5]).

2.1.8 Invasive Weed Optimization

Invasive Weed Optimization (IWO) [6] draws its inspiration from the colonization characteristics of invasive flora as understood from weed biology and ecology. Like ACO, PSO, GA, and SA, IWO is a population-based stochastic algorithm. Weeds exhibit a very strong tendency to opportunistically occupy (colonize) the interstitial spaces is a cropping field. Spaces not occupied by crops, which usually do not spread, become weed-filled, and the weed then grows and propagates by consuming unutilized resources in the field. The weed that uses these resources most effectively becomes the dominant (fittest) weed. When a weed flowers, it produces seeds that are randomly dispersed throughout the field until all interstitial space is occupied and all resources utilized.

Figure 6 shows a flow chart the IWO implementation used to solve electromagnetic problems in [6]. This flowchart starts out much the same as the GA flowchart with a randomly generated population whose fitness is evaluated in the initial step. Each weed then produces a number of seeds (reproduction) based on its fitness, with weeds having better fitnesses being allowed to produce more seeds. The seeds are

then randomly dispersed through the decision space using a normal (Gaussian) distribution of random numbers with mean value equal to the weed's location. After the new seeds have been dispersed, they are allowed to grow into new flowering weeds, and the process is repeated until a convergent solution is generated. Because the number of weeds grows constantly, a maximum weed population serves as a ceiling on weed count. Whenever it is exceeded, the bottom worst plants are "weeded out" by being discarded.

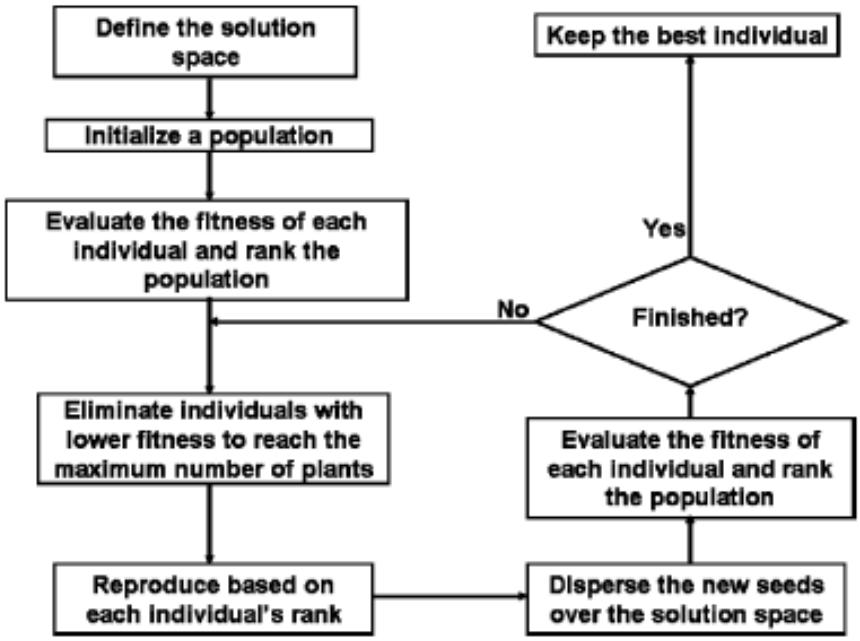

Figure 6. Invasive Weed Optimization flow chart (reproduced from [6]).

2.1.9 Intelligent Water Drop Algorithm

The Intelligent Water Drop algorithm (IWD) [7], like SA and CFO, analogizes a physical process. But, like SA and unlike CFO, it is stochastic in nature instead of deterministic. IWD is inspired by the notion that the seemingly random meanders in a river or stream bed are, in fact, based on mechanisms that can be applied to effectively solve optimization problems. Two principal factors are considered in IWD: water velocity and soil characteristics. Each IWD flows from a source to a destination, initially with non-zero velocity and zero soil. Along its route, velocity and soil both may be gained and lost. Soil inhibits drop velocity, so that between points the IWD's velocity increases inversely with soil (in a non-linear manner). Figure 7 shows typical IWD results for the Travelling Salesman Problem (see also discussion above on SA and TSP).

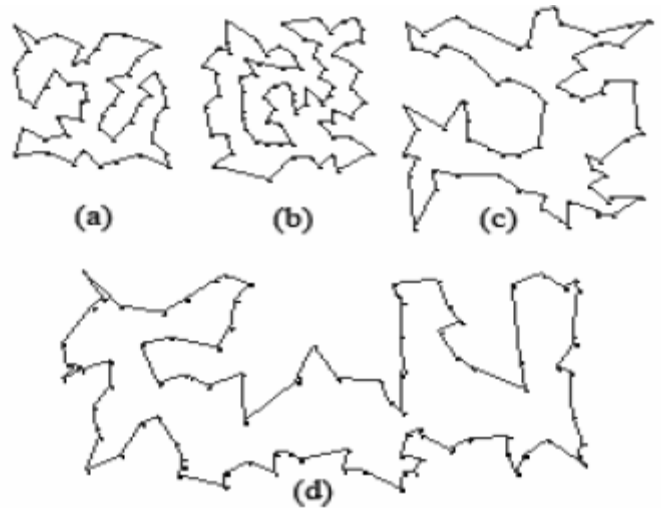

Figure 7. IWD results for the Traveling Salesman Problem (reproduced from [7]).

2.1.10 Bacteria Foraging Optimization

Bacteria Foraging Optimization (BFO) [8] mimics the natural behavior of bacterial seeking food. The motion of individual bacteria is driven by avoiding noxious elements in the environment while "swimming" upward along the food concentration gradient hopefully locating a denser source of food. *Chemotaxis* is the process by which a bacterium tumbles to orient itself, swims a fixed distance, and samples the food concentration at its new location. If the concentration is greater than at the previous location, then the bacterium continues in the same direction for another step. But if the concentration is lower, then the bacterium tumbles into a new direction and explores it instead. Each bacterium has a finite lifetime that limits the number of steps it can take. At the end their lifetime bacteria that are in the highest food concentration regions are allowed to "reproduce" by splitting into two new bacteria, while those in other regions die. Because the bacteria are dispersed, after the reproduction step, only the best half are retained, while the others die, thus preserving the total number of bacteria. The user specifies the maximum number of chemotactic and reproductive steps, maximum swim length, and maximum swims in a given direction.

2.1.11 An Antenna Optimization Example

This section presents a specific example of an optimized antenna design, a Self-structuring antenna (SSA) that can alter its electrical shape in near-real time in response to factors such as a changing environment or a change in mission. Typically these structures comprise wire segments that can be interconnected using control signals from a micro-controller device. An example of a SSA that was optimized using three different optimization algorithms appears in Figure 8 [9]. The antenna in this case is an asymmetric wire array containing 39 switches resulting in 549 *billion* possible antenna configurations.

The specific problem addressed in [9] was whether or not the optimization algorithms would provide better performance than a simple random search for a configuration that met specific performance criteria.

The wire structure was modeled with the Numerical Electromagnetics Code (NEC) following all modeling guidelines with respect to wire segmentation and segment length to diameter and wavelength ratios. The objective was to obtain low VSWR (< 2) values at frequencies of 40, 100 and 400 MHz. The antenna grid measured 0.6 meter square (0.08λ on a side at 40 MHz), which electrically is quite small. Figures 8, 9 and 10 compare the results of a random search to those produced by the three optimization algorithms. Random search performed very poorly at the lowest frequency, while each of the optimization algorithms performed relatively well. The GA produced the "tightest" results at 40 MHz (minimum standard deviation), and achieved the design objective in 60% of the trials. This example shows that real-time-optimized SSAs are within reach for the set-top antenna application at this time, and that fairly economical designs may be possible.

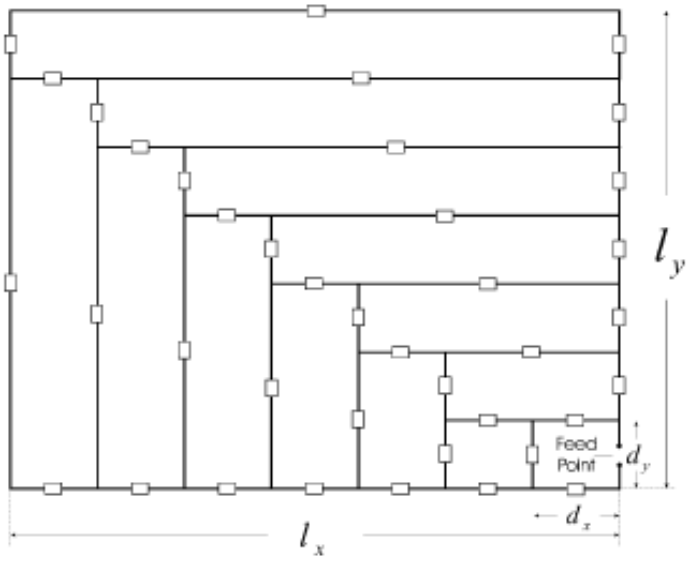

Figure 8. SSA geometry (reproduced from [9]).

TABLE IV
LOWEST SWR FOR A RANDOM SEARCH OF 1500 EVALUATIONS

	400 MHz	100 MHz	40 MHz
Trial 1	1.1157	2.5939	365.4
Trial 2	1.0509	2.5277	958.7
Trial 3	1.0847	1.3926	281.0
Trial 4	1.0251	1.4578	2274.8
Trial 5	1.1516	2.9019	1267.7
Trial 6	1.0936	1.3708	222.7
Trial 7	1.1020	1.5983	28.34
Trial 8	1.0649	1.3312	98.042
Trial 9	1.1206	2.6590	1306.8
Trial 10	1.0816	1.9016	133.4
mean	1.0891	1.9765	694.6
σ	0.0365	0.6317	737.66

TABLE V
LOWEST SWR FOR AN ACO SEARCH OF 1500 EVALUATIONS

	400 MHz	100 MHz	40 MHz
Trial 1	1.0471	1.0246	2.1052
Trial 2	1.0123	1.0371	2.6384
Trial 3	1.0173	1.0418	1.2983
Trial 4	1.0049	1.0363	2.6466
Trial 5	1.0316	1.1012	12.392
Trial 6	1.0075	1.1692	2.6620
Trial 7	1.0314	1.1049	1.7787
Trial 8	1.0366	1.2319	1.3721
Trial 9	1.0046	1.1065	9.9923
Trial 10	1.0106	1.1854	6.9844
mean	1.0204	1.1039	4.3870
σ	0.0151	0.0719	3.9714

Figure 9. SSA VSWR, random search [left], ACO [right] (reproduced from [9]).

TABLE VI
LOWEST SWR FOR AN SA SEARCH OF 1500 EVALUATIONS

	400 MHz	100 MHz	40 MHz
Trial 1	1.1687	1.1516	3.0731
Trial 2	1.0203	1.0225	1.4967
Trial 3	1.0108	1.0644	18.143
Trial 4	1.0028	1.1539	3.2275
Trial 5	1.0019	1.2632	1.6679
Trial 6	1.0071	1.1619	1.6165
Trial 7	1.0014	1.0465	1.1154
Trial 8	1.0105	1.1120	8.3711
Trial 9	1.0029	1.1734	4.2047
Trial 10	1.0246	1.0263	4.5326
mean	1.0251	1.1176	4.7449
σ	0.0511	0.0775	5.1755

TABLE VII
LOWEST SWR FOR A GA SEARCH OF 1500 EVALUATIONS

	400 MHz	100 MHz	40 MHz
Trial 1	1.0096	1.1212	1.0864
Trial 2	1.0092	1.0909	2.8490
Trial 3	1.0096	1.0778	2.2738
Trial 4	1.0057	1.0414	1.4017
Trial 5	1.0086	1.0482	2.3839
Trial 6	1.0180	1.0764	1.9164
Trial 7	1.0041	1.0694	1.2298
Trial 8	1.0126	1.0873	1.7648
Trial 9	1.0056	1.0942	1.7939
Trial 10	1.0032	1.0735	4.9056
mean	1.0086	1.0780	2.1605
σ	0.0044	0.0229	1.1065

Figure 10. SSA VSWR, SA [left], GA [right] (reproduced from [9]).

2.1.12 References

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2.2 Fragmented Antennas

As an additional illustration of the power of an electro-magnetic (EM) optimization algorithm, this section describes the computer program ***FAopt*** that begins with a grid of wires as shown in Figures 1 and 2 and optimizes the placement of these wires using a Binary Genetic Algorithm (BGA) and the NEC-4 [1] Method of Moments (MoM) FORTRAN code. Each bit in the BGA chromosome is either a 1 or a 0, representing the presence or absence of a wire, respectively. Quadrant 1 and 3 are a mirror image of each other as well as quadrant 2 and 4, so that the antenna is symmetrical about its diagonal. The resulting antenna is comprised of a series of plated conductors, some of which are connected to one another. Some are capacitively/inductively coupled and act as parasitic elements to increase the element impedance and radiation pattern bandwidth. These types of radiating elements, when implemented as pixels rather than wires, (see Maloney et al of GTRI [2]) are generally called "fragmented" antennas [3]. ***FAopt*** allows the user to choose the polarization as either vertical or horizontal, and the antenna can be either directional or omni-directional. The user also can choose the dimensions for the desired antenna, as well as the frequency range. This program was used to design a set-top antenna to operate in the VHF and UHF bands. (See Section 3.0). During optimization a window is shown that updates the progress of the optimizer with the best "figure of merit" (FoM) displayed along with the standing wave ratio.

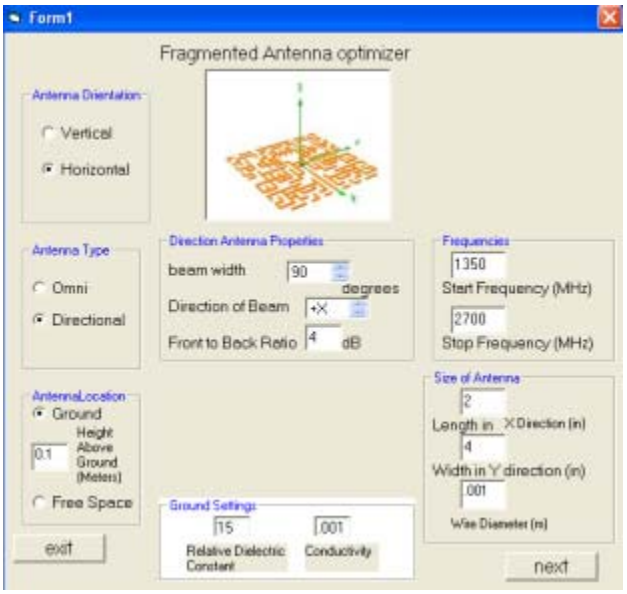

Figure 1: FAopt Code's GUI

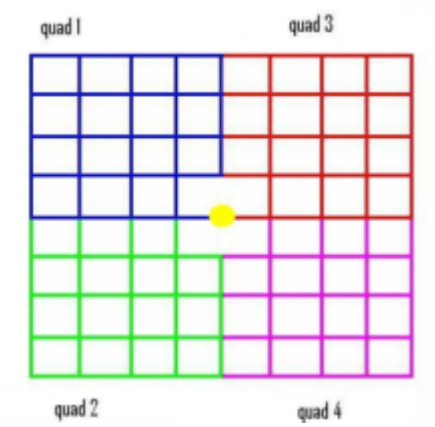

Figure 2: Wire Grid

As an example, a thin planar directional antenna was optimized using ***FAopt*** as a horizontally polarized antenna from 2400-2480MHz with 8 blocks per row/column per quadrant. The orientation appears in Figure 1, and the antenna model is shown in Figure 3. The dimensions for this antenna are 2" by 2". A prototype was built as shown in Figure 4.

Figure 3. Fragmented Directional WiFi Antenna

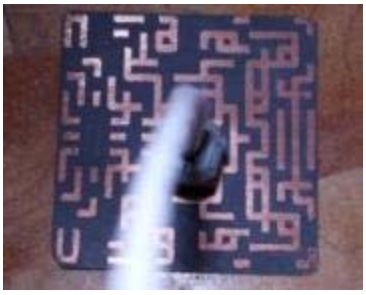

Figure 4. Fragmented Directional WiFi Antenna

The simulation results, which were confirmed by measurements, shows the average gain of this antenna to be better than 1.9 dBi. This antenna is a directional WiFi antenna optimized to have a F/B ratio of about 6 dBi. Its radiation pattern is very consistent

across the band as shown in Figure 6, and its measured VSWR is better than 2:1 across the band. Further work, such as changing the grid template or the optimization routine, could be undertaken to make this approach more efficient for designing set-top VHF/UHF antennas.

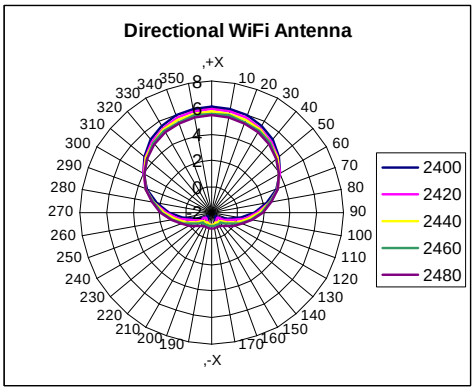

Figure 5. Pattern of Directional WiFi Antenna

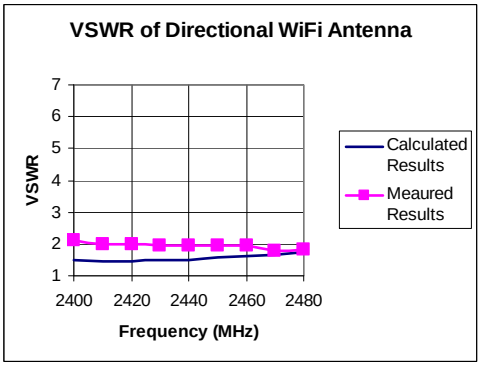

Figure 6. VSWR of Directional WiFi Antenna

The F/B ratio of simulated and measured are in fair agreement as shown in Figure 7.

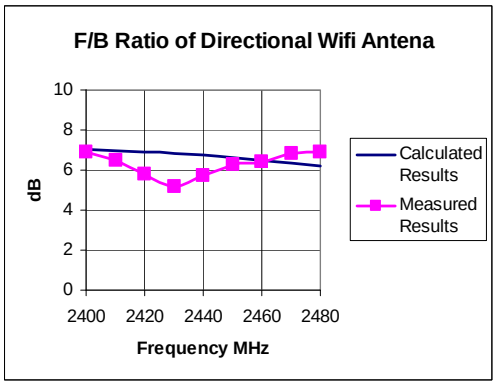

Figure 7. F/B Ratio of Directional WiFi Antenna

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[2] J. G. Maloney, M. P. Kesler, P. H. Harms, G. S. Smith, United States Patent No. US 6323809 B1, Nov 27,2001

[3] B. Thors, H. Steyskal and H. Holter, “Broad-Band Fragmented Aperture Phased Array element Design Using Genetic Algorithms”, IEEE Trans. *Antennas Propag.,* Vol 53, no. 10, pp. 3280-3287

2.3 Non-Foster Impedance Matching

Foster's reactance theorem, which dates to 1924, states that a lossless reactance must have a positive slope with frequency. Any lossless matching network presumably satisfies the theorem, but a great deal of research that has been done on "non-Foster matching" in the past 15 years suggests otherwise. This section examines how non-Foster networks may be useful in set-top antennas to match an electrically small element (characterized by a low real and high negative imaginary complex impedance) within, for example, the low-VHF band.

Until recently no one has created a practical antenna system using this method that showed much of an improvement in performance. This was due to limitations in the semiconductor technology at the time that permitted only narrow band solutions. Semiconductor technology has improved considerably in recent years resulting in increased bandwidth, lowered noise, and decreased losses in the devices. Electrically small antennas present high Q impedances characterized by large reactances and small radiation resistances. For such antennas, the effectiveness of passive matching circuits is severely limited by gain bandwidth theory, which predicts narrow bandwidths and/or poor gain. The result inevitably is a poor signal to noise ratio compared with a larger antenna.

Non-Foster matching uses negative impedance converters (NIC) to create "negative" capacitors and "negative" inductors [1-2]. It is possible, for example, to use negative capacitance to cancel out the reactance in a short dipole or monopole. This leaves only a small real impedance which could then be matched to, say, 75 Ω using a transformer. One disadvantage of this approach is that transformers add to the antenna size which could be undesirable. An alternative to matching could be achieved by placing the non-Foster matching circuit away from the feed to control both the real impedance as well as the reactance of the antenna [3]. The potential improvement in antenna performance is very significant. Theoretical bandwidth improvements have been shown for a small loop (increasing bandwidth from 50 MHz to 300MHz at a center frequency of 700MHz), as well as a small broadband dipole (increasing bandwidth from 250MHz to 1000MHz, lowering the lower operating frequency from 250 MHz to 190 MHz).

It should be noted that the Figures and Tables in this section were taken from references 2 through 5, listed at the end of this section.

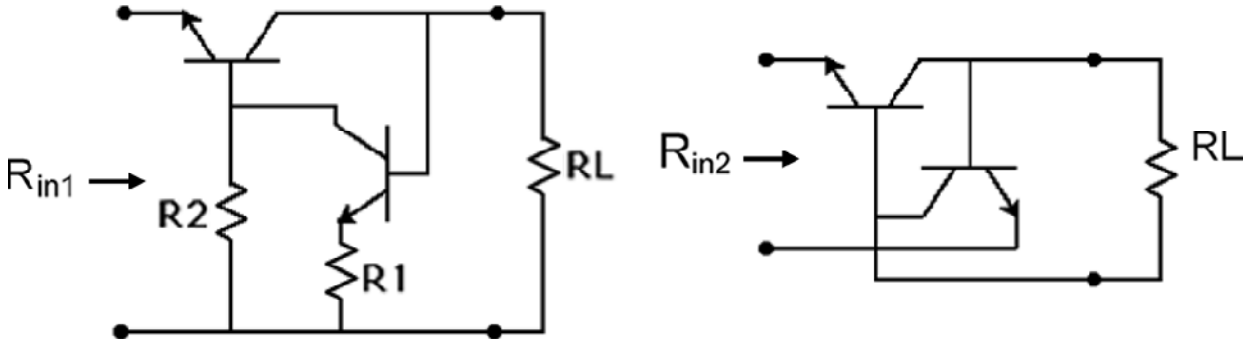
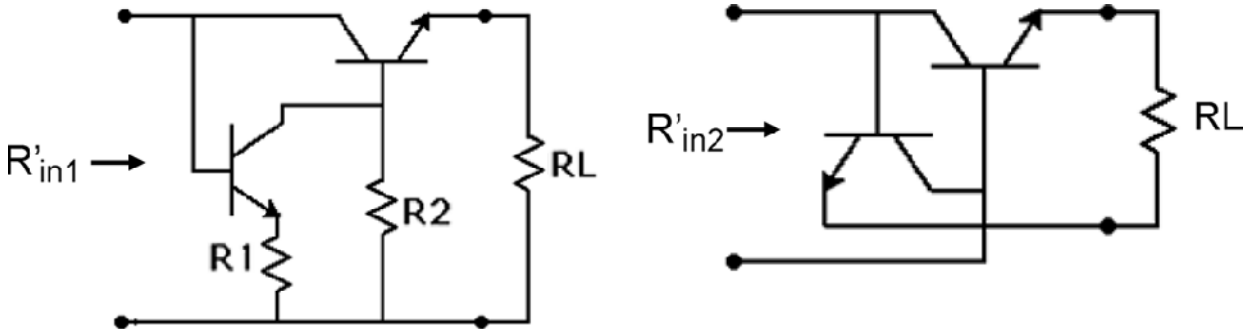


Figure 1. Linvill's ideal OCS NICs terminated in a resistance.



Fiugre 2. Linvill's ideal SCS NICs terminated in a resistance.

With an ideal transistor, a pure negative resistance is achievable. In Linvill's actual realizations (Figures 1 and 2), a substantial reactive component of the input impedance Z accompanies the negative resistance, resulting in a low Q, where the "resistive" Q is defined as: Q=Re(Z)/Im(Z). The circuits shown are "open-circuit stable" (OCS), meaning, practically, that if a very large resistance terminates the negative-resistance one-ports on the left, then the overall network will be stable. The networks of Figure 1 can be turned around as shown in Figure 2, where negative resistances of the "short-circuit stable" (SCS) variety are obtained. Again, practically speaking, this means these networks will be stable if a very large conductance is placed across the input. An initial step in improving the power efficiency of non-Foster matching is simply to use a single-transistor version of the NIC [4].

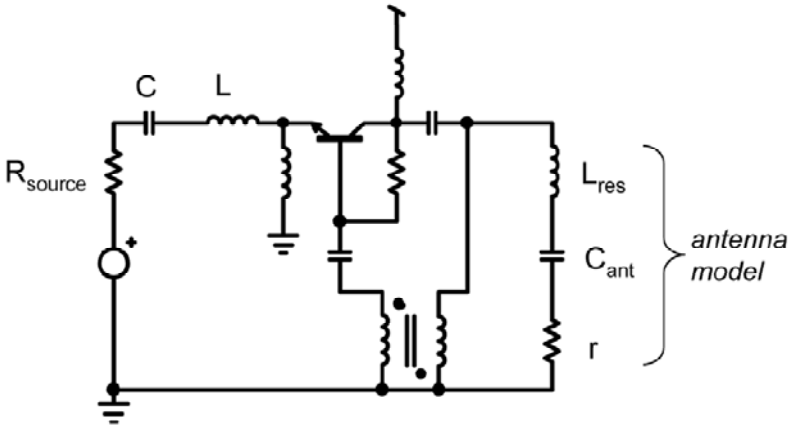


Figure 3. Non Foster Matching

An example of a non-Foster matching circuit is shown in Figure 3. Although not shown in the schematic, a transformer would be used at the input to make the effective value of R_{source}, as seen by the matching circuit, to be 18.75Ω (4:1 impedance ratio).The circuit negates antenna's Capacitance C_{ant} and the voltage-managing resonating inductor L_{res}, as well as the radiation resistance r. The idea is to cancel the negated L_{res} and C_{ant} with C and L on the left, respectively. The negated r is absorbed by the much larger source resistance and does not affect operation of the circuit. The source resistance would be 75 ohms in this case.

An example of non-Foster matching for a 12 inch dipole is shown in Figure 4. It is shown in Figure 5 that this non-Foster matching network creates a negative capacitance that is used to cancel out the reactance of the 12" dipole. This is used along with a transformer to help match the radiation resistance of the antenna to the source impedance of 50 Ω. The non-Foster matching network is compared to a similar dipole with a lossy passive network for improvements in signal to noise as shown in Figures 6 and 7. The improvement in SNR is quite remarkable, as much as 25 dB around 44 MHz.

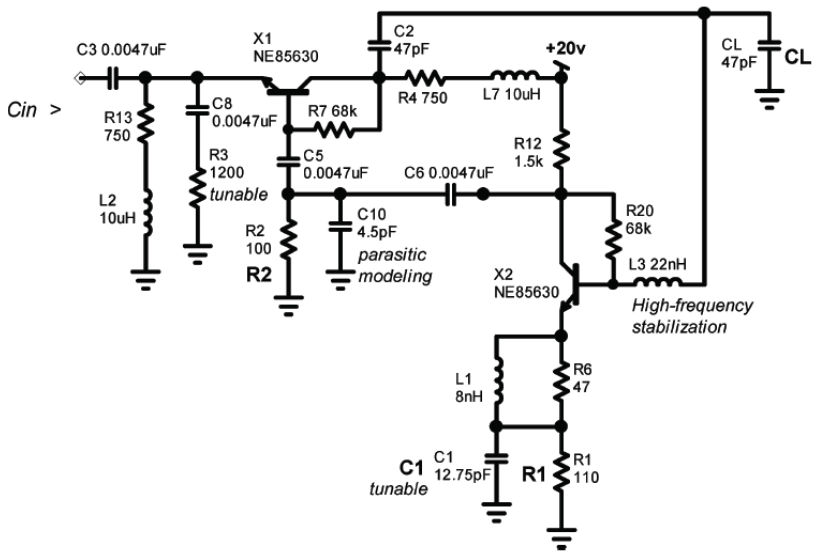


Figure 4. Non-Foster Matching Network for a 12" Dipole.

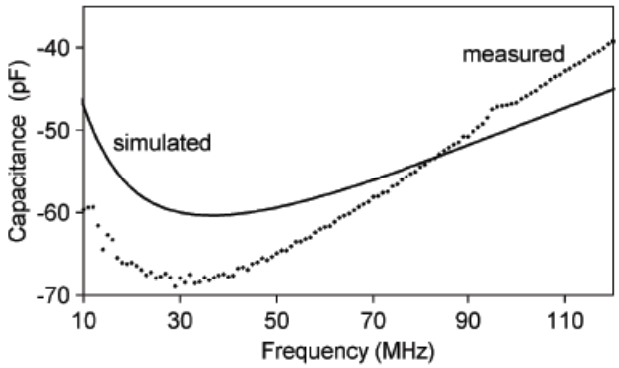


Figure 5. Measured and simulated results for capacitance of non-Foster matching network.

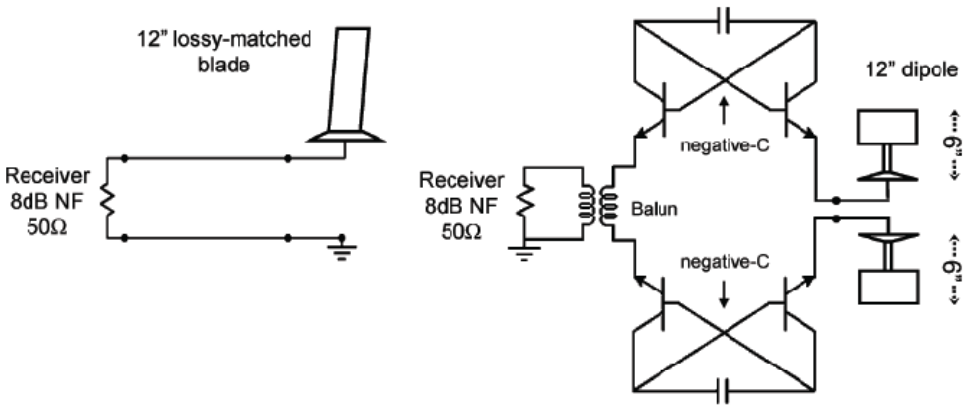


Figure 6. Setup for Signal to Noise improvement measurement.

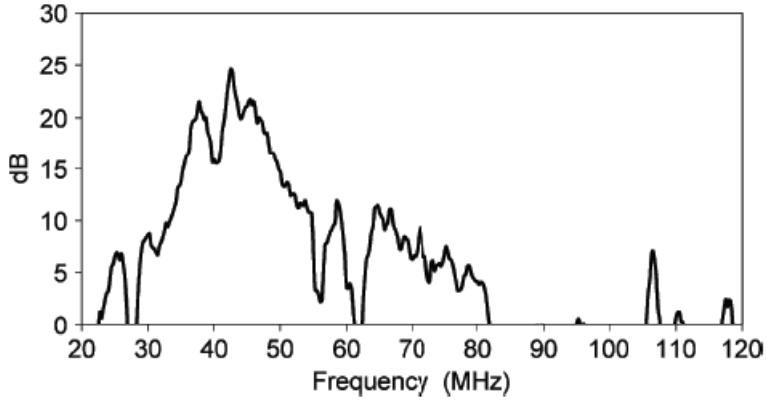


Figure 7. Measured Improvement of Signal to Noise

Non-Foster matching may be particularly useful for the set-top application because it works best with electrically small antennas. Therefore, this type of match could be useful when designing a small VHF/UHF DTV set-top antenna. Improvements in signal to noise measurements have been shown from 20MHz to 400 MHz [5]. In particular, the use of non-Foster techniques to impedance match a lossy electrically-small dipole antenna has been quite effective. On an antenna range, there was a measured 30 dB gain improvement over 60 – 200 MHz, with several dB of improvement as high as 400 MHz. Again, the comparison was to the same antenna with no matching at all. Although no S/N measurements were made, the circuits used were based upon the same low-noise designs developed earlier for lower-frequency circuits. Because of the lossiness of the antenna, passive matching can do a little better than no matching at all, and these results are illustrated in simulation. This computer simulation designed a number of "best-effort" passive matching networks and calculated the transducer gain (S21) between a 50 ohm source and the complex impedance of a 6-inch monopole. S21 for no matching and S21 for a single, ideal negative capacitance whose value (-5.54pF) exactly cancels the antenna reactance at 30 MHz, were calculated. Plots of the real and imaginary parts of the antenna impedance are shown in Figure 8. The various matching networks are illustrated in Figure 9; each of these would appear in place of the NIC in Figure 10b.

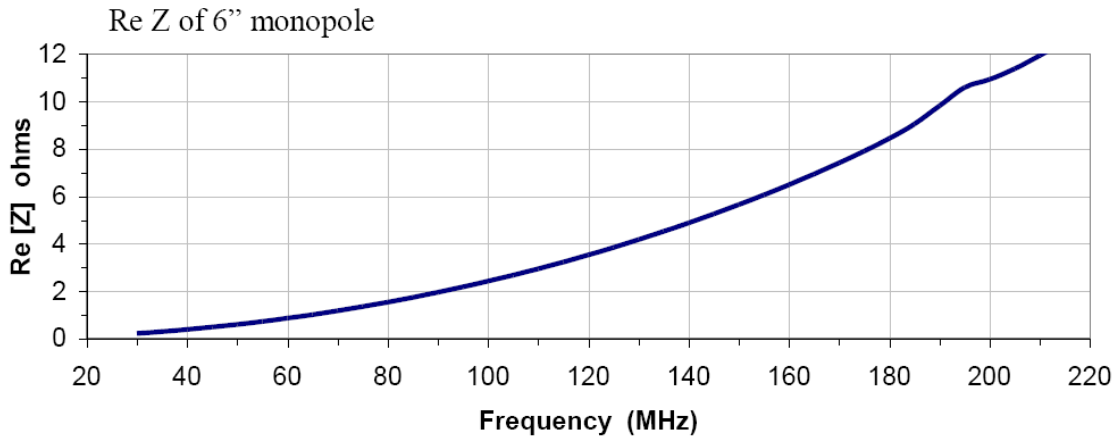


Figure 9-A. Real Part of Antenna Impedance for a six inch monopole

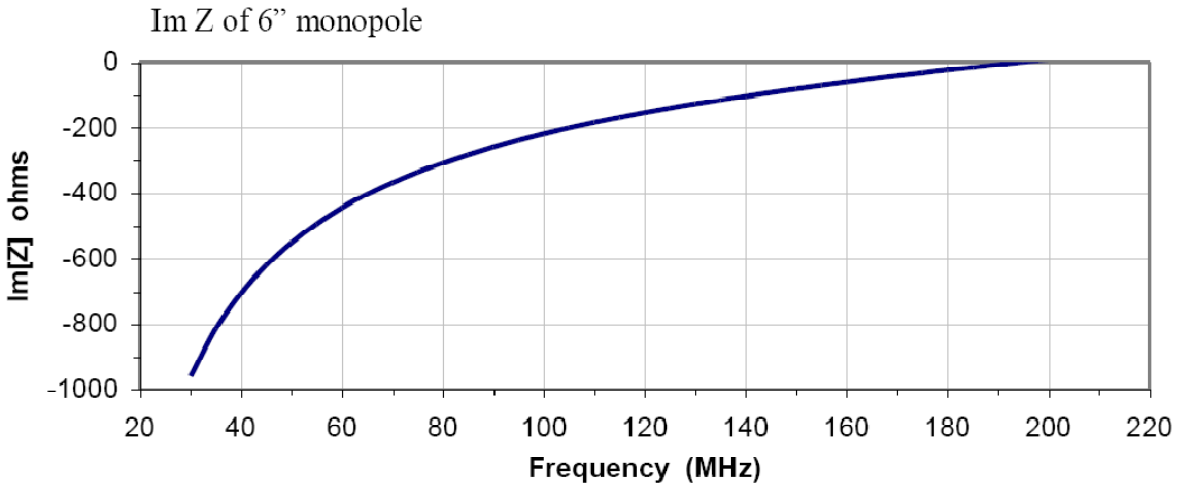


Figure 9-B. Imaginary Part of Antenna Impedance for a six inch monopole

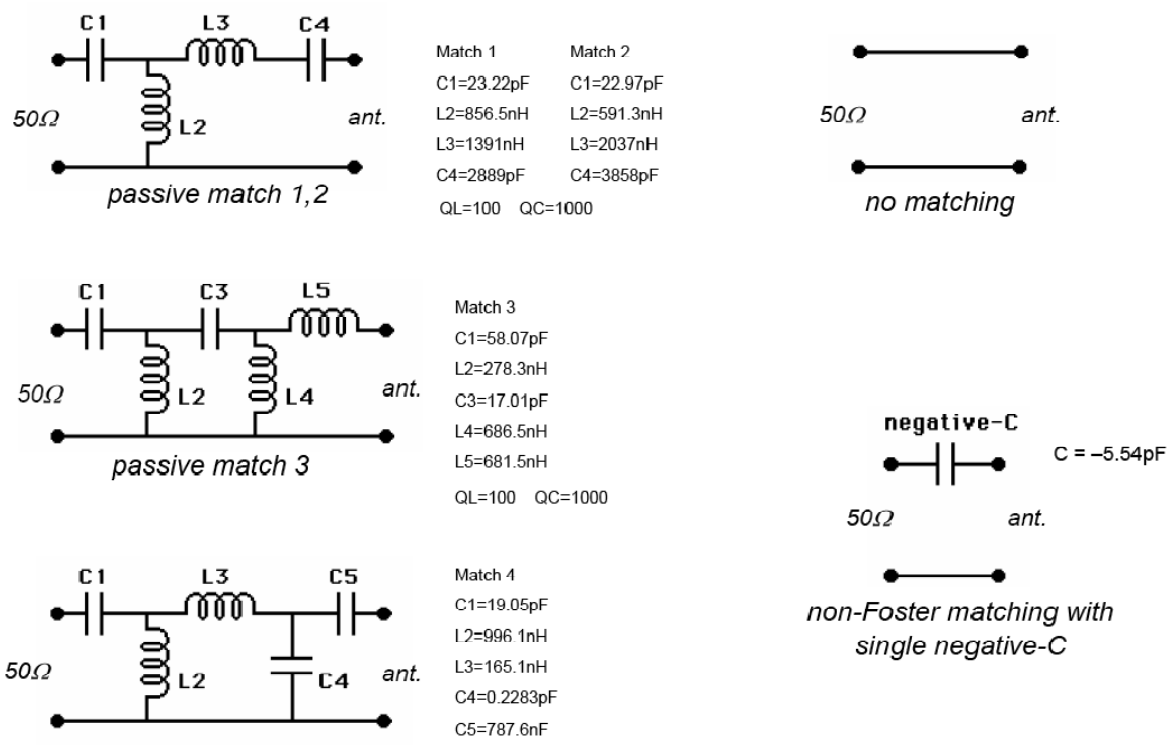


Figure 10. Various Matching Techniques for six inch monopole

Average gain 30-220 MHz	
Non-Foster neg-C:	**-13.2 dB**
Passive match 4:	-21.9 dB
Passive match 3:	-23.1 dB
No matching:	-23.9 dB
Passive match 1:	-26.7 dB
Passive match 2:	-28.4 dB

Table 1. Computed Average Gain of all matching techniques.

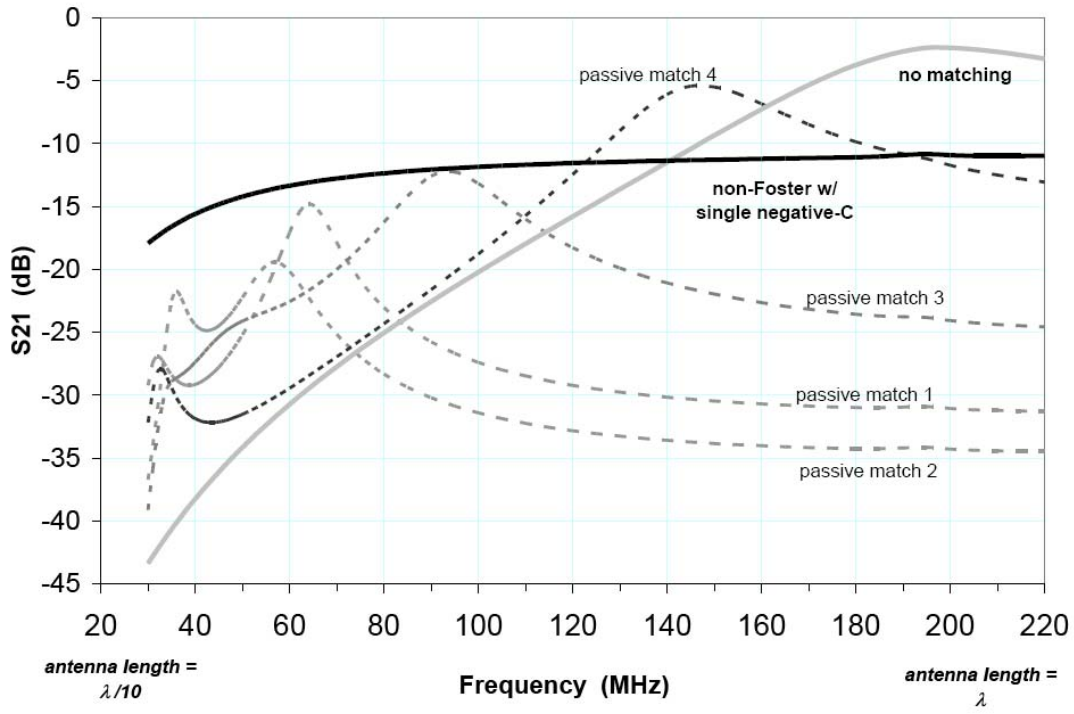


Figure 11. Plots of Computed Gains of all matching techniques for a six inch monopole

The six inch monopole was then built using both the non-Foster matching network and an unmatched antenna. They were both measured with the results shown in Figure 12. In the design, NE68019 BJTs were used in the NICs because these devices provided good overall performance over the entire frequency band. Again, the results are quite remarkable, with the non-Foster antenna showing average gain improvements between about 2 and 28 dB. It is noteworthy that the greatest improvements are at lower frequencies, between about 60 and 210 MHz, approximately the low and high VHF bands.

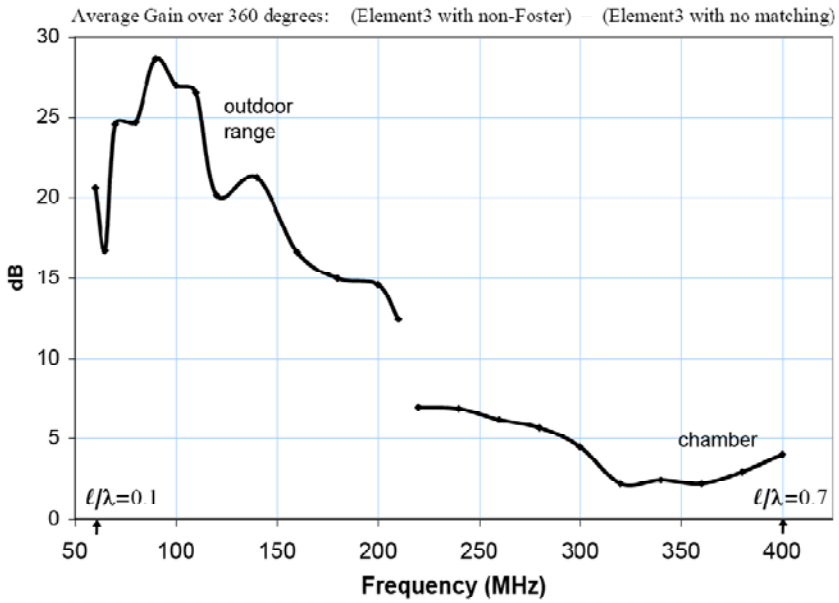


Figure 12. Gain of a non-Foster matching network over the same antenna with no matching.

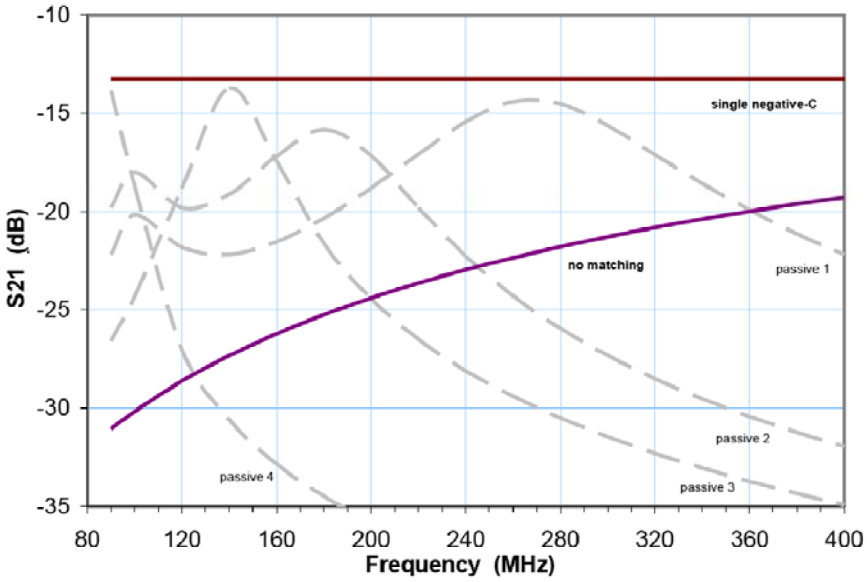


Figure 13. Plots of Computed Gains of all matching techniques for a six inch dipole

Results for an electrically small lossy dipole are shown in Figure 13 (this antenna was simulated, not measured). The actual gain may be less due to noise of the device if this were to be built and tested. Thus, realization of non-Foster technology is still limited by the analog circuitry performance. With respect to the set-top application, it is very reasonable to expect that as better silicon devices are developed covering the entire television frequency range will be possible with substantially better antenna performance in terms of gain and signal-to-noise ratio.

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[2] J. T Aberle, R. Leopsinger-Romak, C. A Balanis, “Antennas with Non-Foster Matching Networks”, Morgan & Claypool Publishers, 2007

[3] S. Koulouridis and J.L. Volakis, “Non-Foster circuits for small broadband antennas”,

[4] S. E. Sussman-Fort and R. M. Rudish “Non-Foster Impedance Matching of Electrically-Small Antennas”,IEEE Transactions on Antennas and Prop., Vol 57,NO. 8, August 2009

[5] S. E. Sussman-Fort and R. M. Rudish, “Non-Foster impedance matching of a lossy, electrically-small antenna over an extended frequency range” presented at the Antenna Applicat. Symp., Allerton Park, IL, Sep. 2007

2.4 Active RF Noise Cancelling

Considerable work has been done to mitigate noise from devices near receiving antennas, as is particularly useful in set-top antennas. An antenna placed on top of a television is especially vulnerable because it may pick up noise from internal circuitry. But this noise can be reduced, which in turn reduces the noise floor, or, equivalently, increases the signal-to-noise ratio (SNR). Intersil [1] has created the QHx220 chip that accomplishes this e.g. at UHF frequencies, or more precisely from 300 MHz up to >3GHz. MegaWave has evaluated the Intersil noise cancelation chip QHx220 shown in Figure 1. The chip was tested for signal to noise improvements at 535MHz as shown in Figure 2. Measured results showed an improvement of about 12dB in SNR as shown in Figures 3 and 4. This technology has been extended down to FM covering the VHF III band [2], which is necessary for this approach to be viable in the set-top environment.

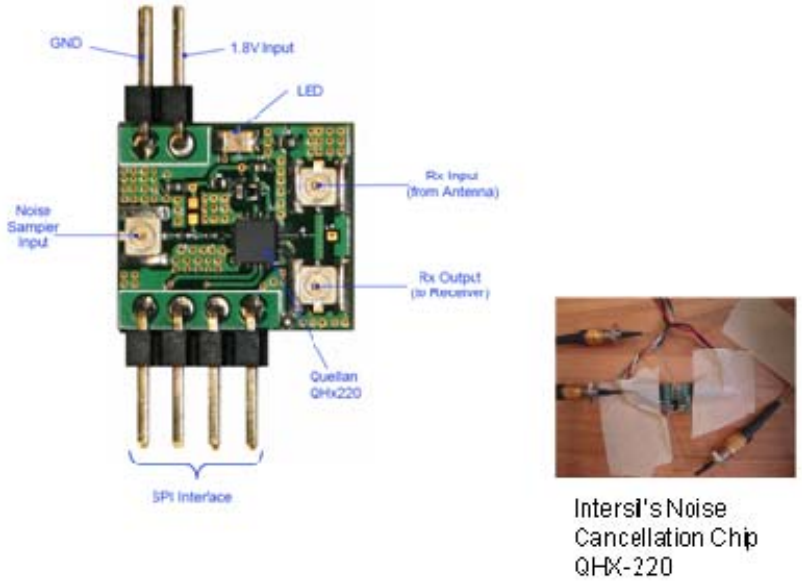


Figure 1: Intersil's Noise Cancellation Chip QHx220

This system works on a principle similar to Bose and Sony noise cancelling headsets, but at a ~1,000,000 times higher frequency. It requires an I and Q setting obtained from the DTV receiver's channel setting and the system's link quality parameter (BER, SNR, RSSI, etc.). This is done on either a micro-processor or in the baseband processor, which run a set of small algorithms. As a result the noise minimum is centered in the desired TV channel and signals - formerly buried in the noise floor - are being restored (Figure 4, e.g. at 549MHz).

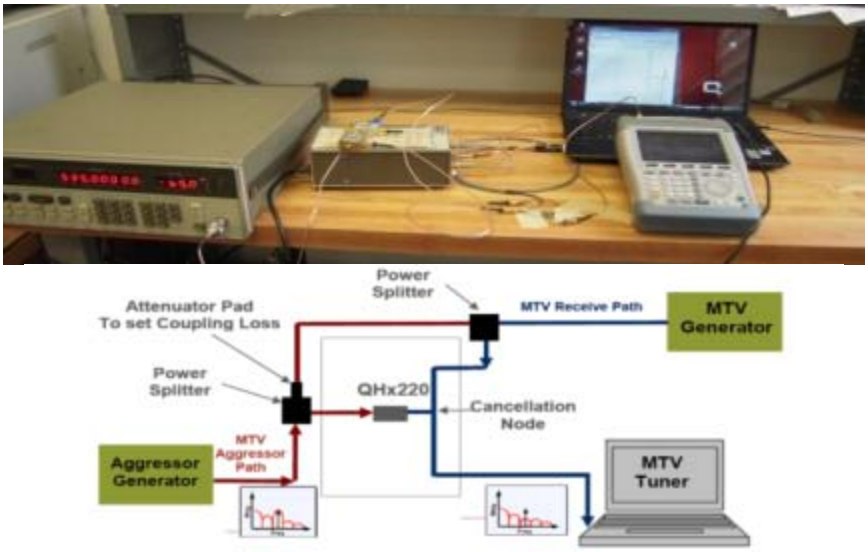
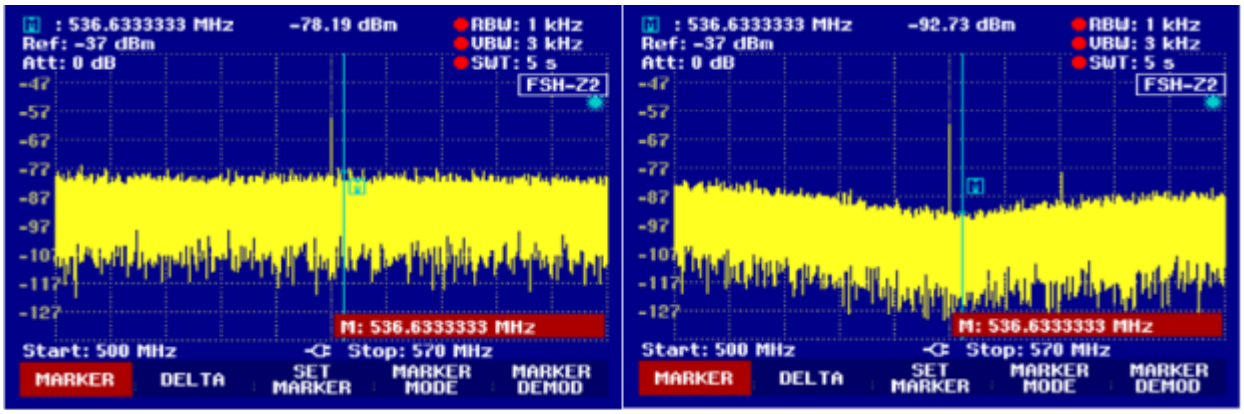


Figure 2 Noise Cancellation Test and Evaluation Setup



Figures 3-4: Measured results before and after noise cancelation

References

[1] Intersil Corporation, 1001 Murphy Ranch Rd, Milpitas, Ca 95035, www.Intersil.com

[2] QHx120 (development) works from 75MHz up to 225MHz.

2.5 Automatic Antenna Matching Systems

2.5.1 Summary

From an impedance matching point of view, the ideal antenna has a perfectly flat, purely resistive input impedance across its entire operating frequency range. The value of that impedance should be 70 Ω because this is the nominal industry standard characteristic impedances of the coaxial cable used in television receivers. Of course, no antenna is perfect. Quite to the contrary, most antennas' impedance variations with frequency are usually quite dramatic. This is particularly true for electrically small antennas, which tend to exhibit low radiation resistances and very high reactances in a narrow frequency range. Set-top DTV television antennas are not necessarily electrically small in the high VHF and UHF bands, but are at low VHF. Due to their required bandwidth (54-698 MHz) they invariably exhibit wide impedance fluctuations.

An impedance mismatch leads to losses and reflected energy that is not transferred to the receiver. These inefficiencies can be mitigated to a great degree by precise impedance matching that is provided by an "antenna tuner" (AT) connected between the antenna output terminals and the receiver input port. The AT may comprise a network of discrete components, or it may be a distributed, continuously varying structure (a tapered transmission line, for example), or some combination of both.

Lumped element matching networks go back to the earliest days of radio, and they often require manual adjustment of the matching elements (usually variable capacitors or roller inductors). In the early 1960s the military developed "ALE" (automatic link establishment) systems that employed remotely controlled automatic antenna tuners. These networks are usually located at the antenna feed and remotely tuned automatically based on VSWR (voltage standing wave ratio) measurements made at the transmitter. Essentially the same approach is taken with the built-in automatic ATs that are common in modern transmitters (most amateur radios contain built-in auto ATs, for example).

Modern low-power ATs use relays to switch lumped elements (inductors, capacitors) in and out of the matching network until the desired match is achieved. This technology is highly developed and readily available for use in set-top TV receive antennas. It therefore is not reviewed in this report. Instead, the emphasis here is on new, developing technologies that may be useful for antenna impedance matching. Four broad categories have been identified.

(a) *Voltage controlled reactive elements* are capacitors and inductors whose values are controlled by an applied voltage. Instead of switching discrete components

in and out of a matching network using relays, the same operation is accomplished by varying the voltage applied to the reactive element. This type of non-mechanical system provides faster response time and more continuous control in a less noisy and probably smaller space than a relay-based AT.

(b) *Integrated ATs* are "antenna tuners on a chip," a complete device that is fabricated on a single chip or in a very small module with only a few external components. Integrated ATs have received considerable attention for cellular applications, and they work quite well. Chip-level devices successfully match up to 10:1 VSWRs in the highly fluctuating cellular antenna environment with rapid response times and high efficiency. These devices are constantly improving, and it is likely that existing technology can meet set-top AT requirements.

(c) *Tunable tracking filters* are already used across the VHF/UHF TV bands for wideband DTV (digital television) tuners, and they work quite well, providing very flat passbands, steep skirts, and harmonic suppression from the tuner's local oscillator. Extremely small devices have been designed and computer-simulated, and some measured data are available confirming achievement of design objectives. This is an evolving technology that may be transferrable to set-top AT requirements.

(d) *Software defined radio (SDR)* is a new concept in wireless communications that seeks to move system properties from the hardware layer to a high-level software layer by developing hardware modules that are fully software-controllable and configurable. SDR radios are gaining wide acceptance in many applications, and it is a developing technology. SDR "smart antenna" technology, which may be applicable to the set-top environment, consequently is a candidate technology for a television AT.

2.5.1.1 **Voltage Controlled Reactive Elements**

(a) *BZN (pyrochlore bismuth zinc niobate)* is a non-ferroelectric dielectric material that shows great promise for a new class of voltage tunable thin film capacitors [1]. When it was introduced in 2006, BZN exhibited the lowest loss of any room temperature complex oxide film with a very high dielectric constant and high tunability of (loss tangent ~0.0005, dielectric constant ~180, ~55% tuning range @ 1 MHz). Losses increase in the microwave range, but are controllable by modifying the electrode structure.

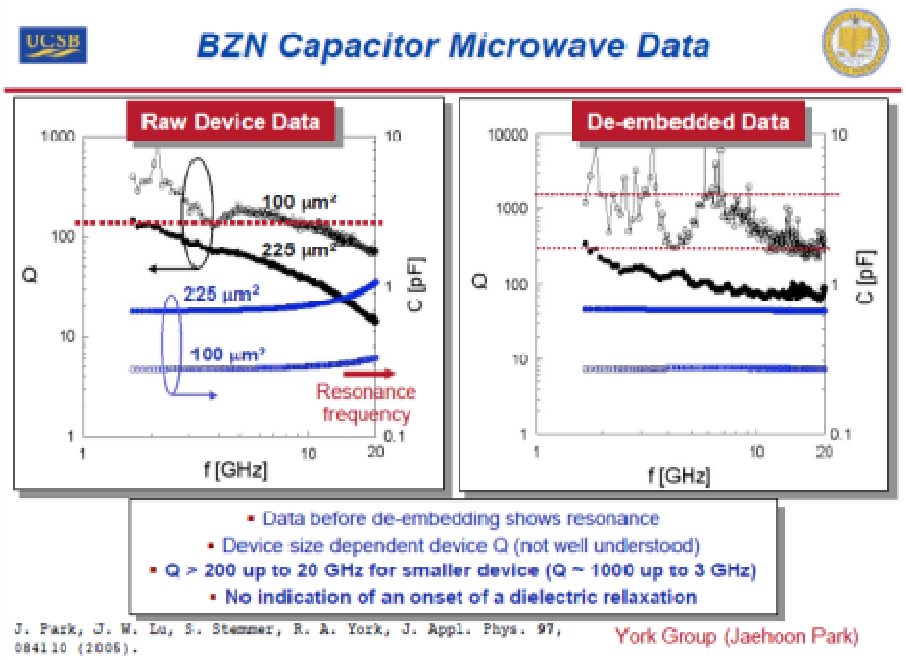


Figure 1. BZN capacitor high-frequency measurements (reproduced from [1])

Measured data for different size BZN thin film capacitors at higher frequencies appear in Figure 1. For small devices (100 μm^2) the *Q*-factor exceeds 200 up to 20 GHz (de-embedded data, right plot). Through about 3 GHz *Q* ~ 1000. In addition, self-resonance in the BZN thin film structure is well above 20 GHz. By comparison, the best state-of-the-art thin-film BST (barium strontium titanate) devices in 2006 exhibited *Q*s that decreased monotonically from about 100 at 1 GHz to around 20 at 20 GHz. BZN is a far superior tunable dielectric. BZN and similar materials (see below) hold considerable promise for the development of completely solid-state ATs that can be used to tune set-top TV broadcast band antennas.

(b) *PZN (lead zinc niobate)* thin film interdigital capacitors (IDC) have been fabricated to increase tunability and reduce bias voltage compared to BZN IDC implementations. Cubic pyrochlore PZN thin film dielectrics provide superior performance through the microwave range. Figure 2 shows a typical configuration. The IDC is fabricated on a silicon substrate in the usual coplanar waveguide (CPW) configuration. Figure 3 plots the measured low-frequency dielectric constant (blue) which increases from just over 200 to about 230 at 10 MHz. The tenability as a function of applied bias voltage decreases from unity to about 0.75 with an applied electric field of 2 mV/cm (bias voltage of 5.5 volts). At 1 GHz the device *Q* falls to about 10 at all bias voltages, but the tunability remains at 26% (compared to 25% at 1 MHz) for 5.5 volts applied bias. The PZN tunable capacitor thus is a strong candidate for voltage-tunable

set-top television antenna ATs, and its performance is comparable or better than that of BZN devices.

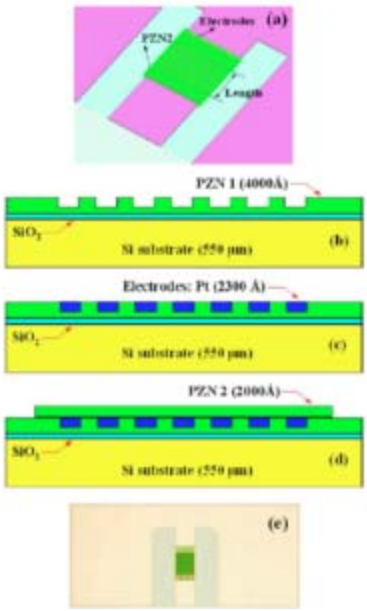


Figure 2. PZN IDC structure (reproduced from [2]).

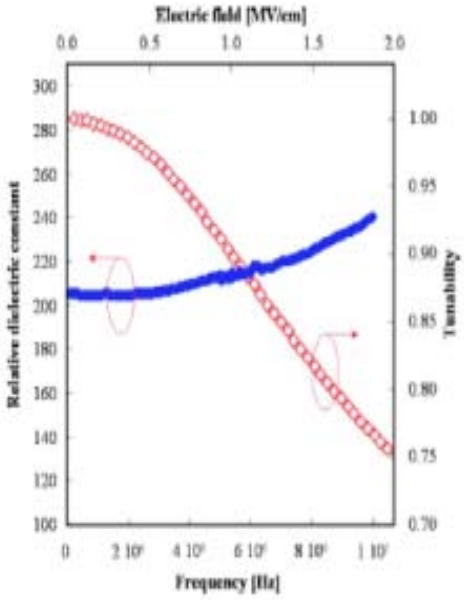


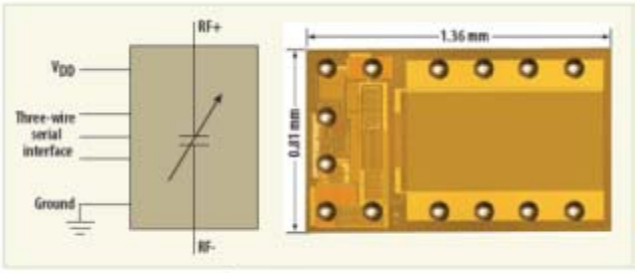
Figure 3. PZN IDC performance (reproduced from [2]).

(c) *DuNE™ capacitors* are a new patent-pending technology developed by Peregrine Semiconductor [3] for DTCs (Digitally Tunable Capacitors). The manufacturing process is based on Peregrine's proprietary UltraCMOS™ technology supplemented by its HaRP™ design methodology. UltraCMOS™ is a patented low-power SOI (silicon on insulator) variant chip architecture that reduces transistor capacitance resulting in increased switching speed. HaRP™ is a design technology that produces significant improvements in harmonic results, linearity and overall RF performance.

The company claims that its DTCs provide digitally controllable capacitances in the range 0.5 to 10 pF with either 3:1 or 6:1 tuning ratios and 32 addressable states (5 bits). *Q*s range from 40 to 80 between 1 and 2 GHz with device switching times below 5 µsec. Power handling capability is quite good (+38 dBm @ 50 Ω), while power consumption is low (100 µamps @ 2.4-3.5 volts DC). The manufacturer claims its DTCs provide better performance than currently available MEMS (*microelectromechanical system*) or BST devices, and they are readily available as off-the-shelf components.

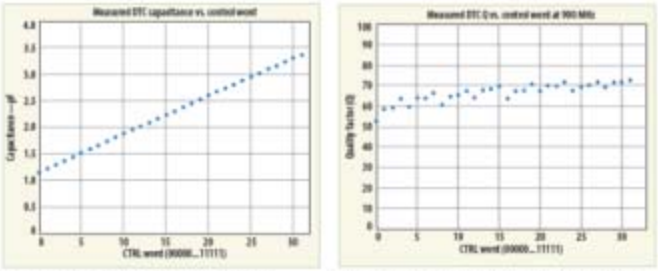
The digitally tuned capacitor schematic diagram and its "flip-chip" package appear in Figure 4. The DTC acts as a series capacitor with two RF terminals. It is powered by a single DC line, and controlled by a three-wire serial interface. The control word is five bits long. The device is extremely small (1.26 mm x 0.81 mm) and thus well-

suited for use in TV set-top antennas. Measured performance is shown in Figure 5. Capacitance variation is very linear with control state (32 increments), and the variation ratio is 3:1, from 1.15 pF to 3.4 pF. The DTC *Q*-factor at 900 MHz, a representative cellular frequency, varies from about 53 to about 72, which is quite good.



5. The schematic diagram shows the DuNE DTC (left) while the photograph (right) shows a flip-chip DuNE DTC prototype that measures just 1.36 x 0.81 mm.

Figure 4. Digitally Tuned Capacitor (reproduced from [4]).



6. The measured performance of a 5-b DuNE DTC prototype shows linear tuning characteristic and a 3:1 capacitance range from 1.15 to 3.4 pF.

7. These measurements show the DuNE DTC prototype with quality factor (Q) designed to be in the range of 60 to 70 at 900 MHz.

Figure 5. DTC measured performance (reproduced from [4]).

DTCs have been effectively applied in cellular phone ATs. A typical configuration is shown in Figure 6. The transceiver, in this case a cellular handset, is connect to its antenna through an AT comprising four main elements: (1) serial interface; (2) digital mismatch sensor; (3) tuning algorithm; and (4) DTC core. This structure is representative of ATs based on other tuning technologies besides DTCs. For example, this same architecture would be used with MEMS-based or tunable thin film capacitor ATs. Note that the diagram does not show the RF path between the antenna and transceiver, only the AT control data path. The serial interface exchanges digital data with the transceiver on the AT status. The mismatch sensor responds to reflected power resulting from any impedance mismatch between the antenna and the system's characteristic impedance. Digital mismatch level data are processed by the tuning algorithm that determines which DTCs should be activated and at what capacitance level. Control words are sent to the DTCs in the matching network (DTC core) to effect the impedance match between the antenna and transceiver.

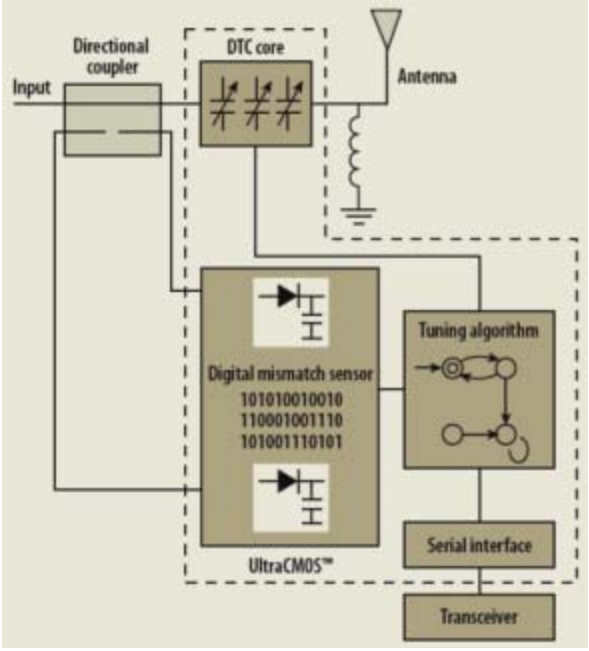


Figure 6. DTC-based AT (reproduced from [4]).

(d) *Dual-gap Tunable MEMS Capacitors* are made using a novel fabrication process that creates two gap device that has an extremely wide tuning range (as much as a factor of 15) with high *Q* [5]. Figure 1 shows the dual-gap structure schematically and as implemented using a "two hump" sacrificial layer which, when removed, creates two MEMS gaps instead of one. In the schematic, the two moveable electrodes are indicated by the double downward arrows. As these suspended elements move closer to the substrate electrodes the capacitance increases. The structure is a MEMS device designed for a linear capacitance variation with applied control voltage.

In the fabricated prototype, the central region (E_c) is 250µm x 80µm (LxW) with a 1.5µm fixed gap. The two "beams" are 800µm long by 80µm wide creating a variable gap because they are movable. The "actuation area" patches (E_a) are 200µm long and 80 um wide with a gap of 4.5µm with the beams not deflected. This structure produces a minimum capacitance of 0.12 pF that is voltage controllable to a maximum capacitance of 1.77 pF. The resulting tuning range is $\frac{C_{\max} - C_0}{C_0} = 1375\%$, which is quite remarkable. The required bias ("pull in") voltage is less than 12 volts. This device is not commercially available, but it represents a class of tunable devices that should be very useful in the set-top AT when (and if) they become COTS (commercial off-the-shelf) items.

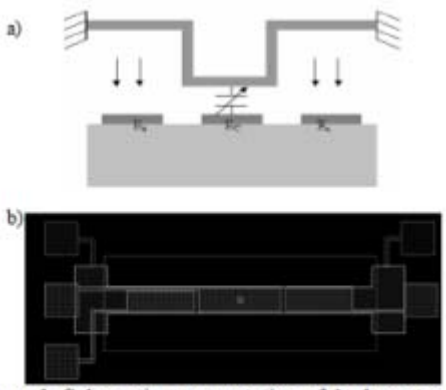


Figure 1: Schematic representation of dual gap tuneable capacitor - a) Cross section - b) Layout top view.

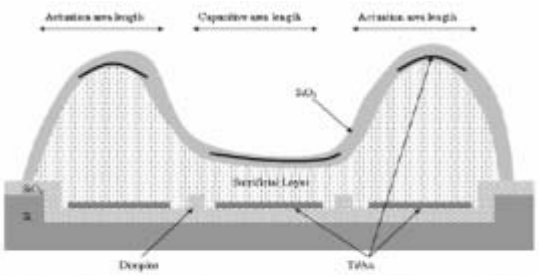


Figure 2 : Schematic representation of the tuneable capacitor based on dual pick profile of the sacrificial layer

Figure 7. Dual-gap tunable MEMS capacitor (reproduced from [5]).

(e) *MEMS floating dielectric capacitors* (MFDCs) are a recent promising development [6]. The movable dielectric is a new actuation principle in which a floating movable dielectric is electrostatically maneuvered to vary the capacitance. Figure 8 shows the new concept schematically. A mechanical spring returns the movable dielectric to its undisturbed position when no force is applied by the electrostatic comb drive. When the dive is activated, the dielectric moves closer to the top and bottom capacitor plates thereby increasing the capacitance. Preferentially the RF path is through the plates, not through the dielectric and spring, which increases losses.

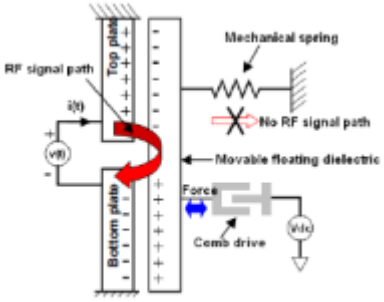


Fig. 1: Conceptual schematic view of the proposed tunable capacitor with an electrically floating dielectric.

Figure 8. MFDC concept (reproduced from [6]).

Figure 9 shows schematic RF signal path superimposed on the MFDC's actual structure as fabricated. In the left pane the signal flows through the spring, which increases the capacitance at the expense of increased losses. In the right pane the signal flows between the top and bottom plates through the movable dielectric. This configuration results in lower losses, but lower capacitance as well. Of course, either configuration can be used, depending upon the specific application.

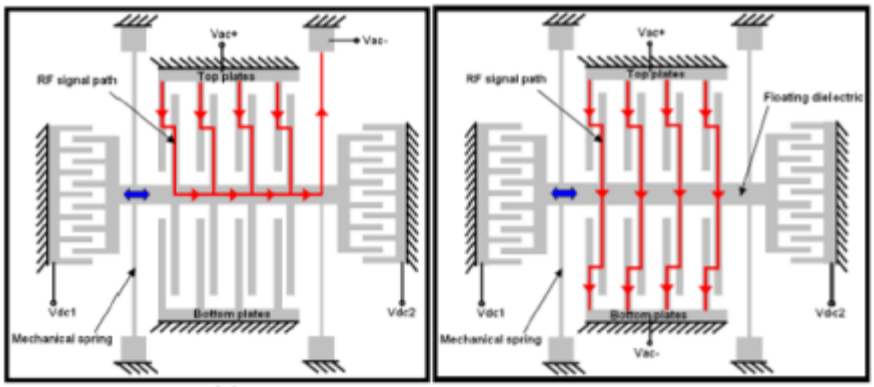


Fig. 2: Schematic views of tunable capacitor design with RF signal through spring (a) and dielectric (b).

Figure 9. MFDC as fabricated and measured data (reproduced from [6]).

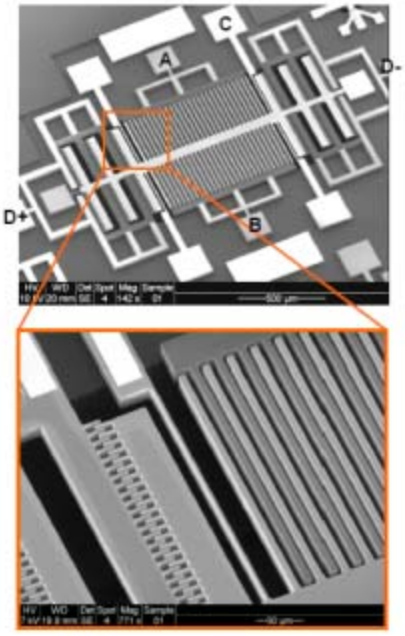


Fig. 4: SEM image of the micromachined tunable capacitor.

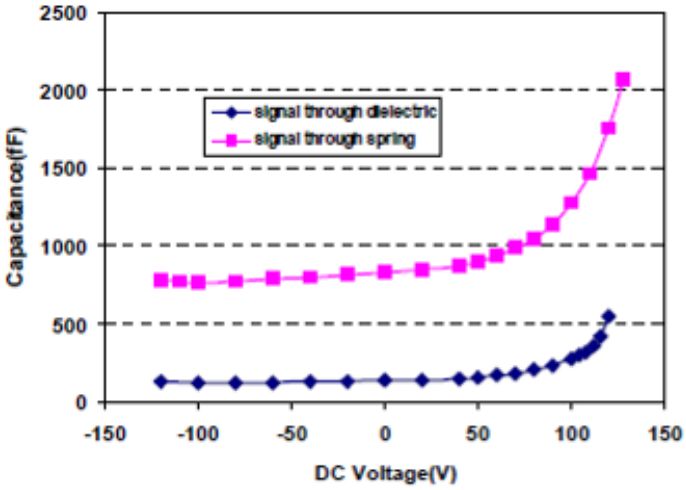


Fig. 5: Measured capacitor changes vs. applied dc bias

Figure 10. MFDC as fabricated and measured data (reproduced from [6]).

Figure 10 provides a scanning electron (SEM) microscope image of a fabricated MFDC (the area outlined in orange, lower pane) and measured performance data. Comparing Figs. 9 and 10, the features in the schematic are readily identifiable in the SEM photo. The actuator pads are labeled D+ and D- in the photo, while the spring is attached to pad C. Pads A and B are the connections for their associated comb plates.

The applied bias voltage ranged from -120 VDC to +120VDC, resulting in capacitance ranges of about 760 fF (femtoFarad) to 2100 fF for RF signals passing through the spring element (pink curve). The initial capacitance (no bias voltage) was 830 fF, leading to a tuning range of approximately 170%. The 1 GHz *Q*-factor was 0.35, which is low by comparison to other technologies. When the RF path is through the movable dielectric (blue curve), capacitance ranged from just over 100 fF to about 550 fF yielding a tuning range of approximately 370% with an initial capacitance of 135 fF. As expected, the reduced loss resulting from routing the signal away from the spring results is a much better *Q*-factor of 56 at 1 GHz. Overall, this performance is not as good as that provided by other technologies, but MFDCs are a new concept that requires further development. The MFDC approach certainly merits consideration as a potentially useful future technology for the set-top AT.

(f) The v*oltage controlled semiconductor inductor* (VCSI) is another reactive component that should be useful in set-top ATs. Inductor values are usually fixed, so that obtaining specific value of inductance in an AT is usually accomplished by mechanically switching small inductors in and out of the matching circuit using relays or MEMS RF switches. Recently disclosed VCSI devices [7] should be particularly useful at broadcast TV frequencies. The device addresses the problem of limited tuning range provided by voltage tunable capacitors by varying the inductance instead.

Generally is comprises regular coil turns of wire interconnected by semiconductor diodes that can connect individual turn to create a VCSI. Including a resistor and capacitor creates a complete tunable RLC circuit. Figure 11 illustrates this patented technology. The left pane shows a perspective view of the device, which includes conductive loops (205) connected to a semiconductor bar comprising P- and N-type regions at each end (208 and 210, respectively) connected by a depletion region (212). Applying a voltage across terminals L_A and L_B varies the length of the depletion region which acts as an insulator. Because its size is proportional to the applied bias voltage, individual coil turns are either connected or disconnected in proportion to the voltage, thereby creating a voltage-variable inductor.

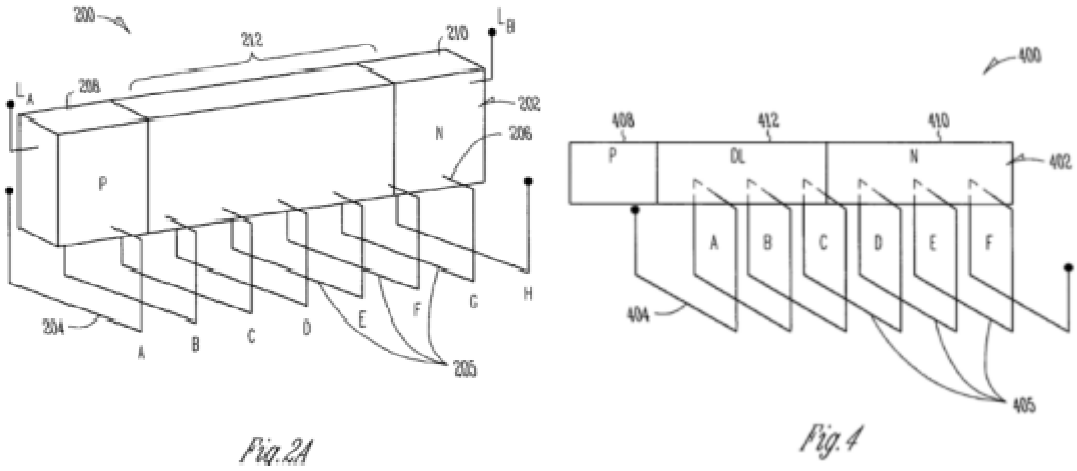


Figure 11. Voltage controlled semiconductor inductor (reproduced from [7]).

A helical-turn implementation of the VCSI is shown in the right pane of Figure 11. The same semiconductor diode structure comprising P- and N-type end regions connected to a central depletion layer (DL) whose length is voltage-variable. The figure provides a schematic representation of how the three regions' lengths vary with bias voltage compared to the "off" state shown in the left pane. In the diagram, coil turns C through F are electrically short-circuited because they are outside the insulating DL. In this case, only turns A and B are active elements in the inductor.

The VCSI may become an important element in set-top ATs because it provides a complete RLC tuning in a single chip-level device. A bank of VCSIs, for example, could comprise the switchable reactive elements in a set-top AT that are completely voltage-controllable, thereby eliminating the need for mechanical switching relays. VCSI therefore is an attractive emerging technology that merits watching.

2.5.1.2 Fully Integrated ATs

In addition to the emerging component-level technologies described in §1, fully integrated ATs have been developed that also merit consideration for set-top tuners. This section examines developments in that area.

(a) A *reconfigurable RF-MEMS-based matching network* is described in [8]. The chip-level device's circuit diagram is shown in Figure 12. It comprises two stages, the first of which is a Pi-match section with four shunt capacitor-series inductor (CL) sections. A total of eight RF-MEMS switches are employed yielding 2^8 impedances. The shunt capacitors are formed from bi-valued MEMS varactors. The variable capacitor in series with each fixed inductor, also a MEMS varactor, has the effect of adjusting the series inductance. The second section is a phase-shifter comprising a 3-dB 90-degree coupler connected to a reflective load. MEMS varactors adjust the load reflection

coefficient to control the overall phase shift. A total of 2^3 phase rotations are possible in the phase shifter, resulting in as much as 340 degrees of total phase shift that can be applied to the impedances at the output of the Pi-match. The network is designed to work at a 50 Ω impedance level.

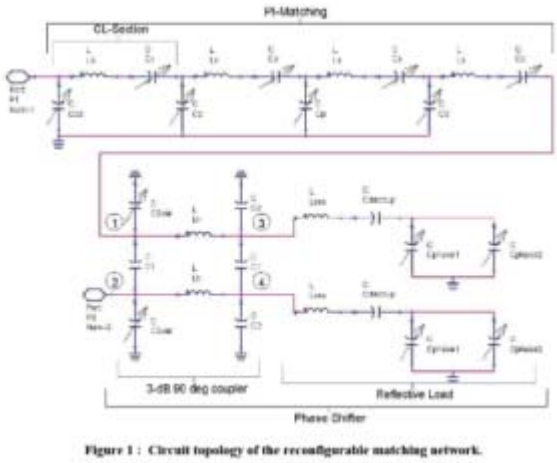


Figure 12. Reconfigurable RF-MEMS-based matching network circuit (reproduced from [8]).

The chip layout and fabricated device are shown in Figure 13 with the major sections being labeled on the layout diagram. The chip area is extremely small, only 40 mm^2 (slightly larger than 3x12 mm). The published report indicates that the chip has been fabricated and was undergoing testing, but no actual measured data were reported. Instead simulated performance at a single frequency (620 MHz) was calculated at each of the possible 2,048 impedance combinations and plotted on a Smith chart as shown in Figure 14. The circles are impedance values computed for the first stage alone, while the dots represent the show the performance of the complete reconfigurable matching circuit. The fairly uniform distribution of dots throughout the Smith chart suggest that the matching circuit will effectively match an extremely wide range of impedances to 50 Ω. This technology is directly applicable to the set-top AT and appears to be on the verge of realization. Future published results for reconfigurable RF-MEMS-based matching networks clearly bear watching.

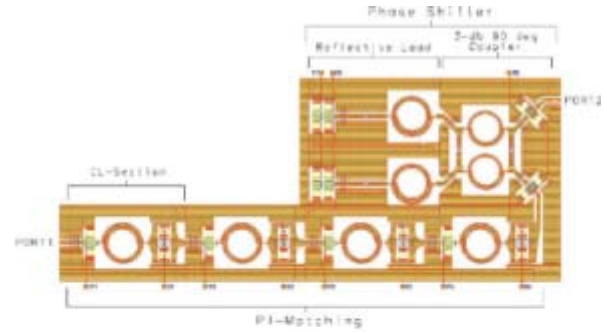


Figure 3 : Layout of the circuit as implemented in ADS

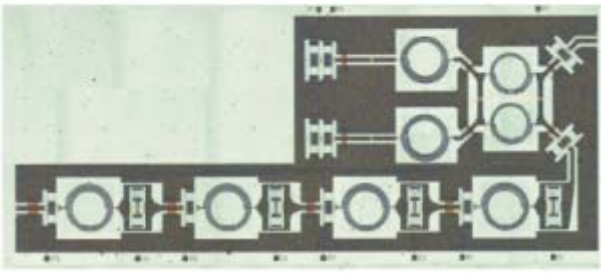


Figure 4 : Photograph of the fabricated circuit

Figure 13. Reconfigurable RF-MEMS-based matching network layout & fabrication (reproduced from [8]).

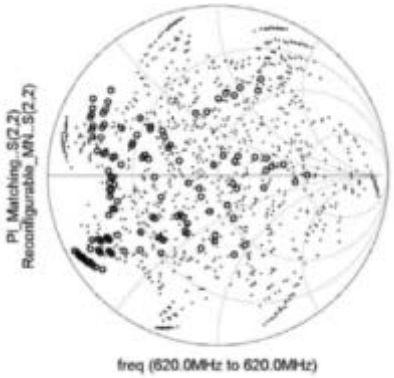


Figure 2 : Simulated impedance values (circles: first stage; dots: full reconfigurable matching network).

Figure 14. Reconfigurable RF-MEMS-based matching network performance (reproduced from [8]).

(b) The *AT using only RF signal amplitudes* described in [9] may be useful for the set-top application because it specifically addresses the issue of highly variable, unpredictable and uncontrollable environments. This new technology was developed for mobile applications such as cellular transceivers because their fluctuating environment often causes VSWR spikes approaching 10:1 in nominal 50 Ω systems. These fluctuations are frequently transient on time scales in the millisecond range. The new AT design accommodates this environment, and it may be useful in the highly variable TV set-top antenna environment as well.

The architecture appears in Figure 15, which includes schematic Smith chart representations of the impedance at various points. ATs generally comprise an impedance sensor, a tunable matching network, and control circuitry that changes the network parameters in order to achieve an acceptable VSWR (usually relative to 50 Ω with VSWR < 2-3:1). The general approach shown in Figure 15 therefore is applicable to any matching network. The new matching concept involves a two-step process: (1) the reactance is essentially tuned out using a series or shunt reactance; and (2) a tunable "transformer" changes the remaining resistive component to the desired value of impedance.

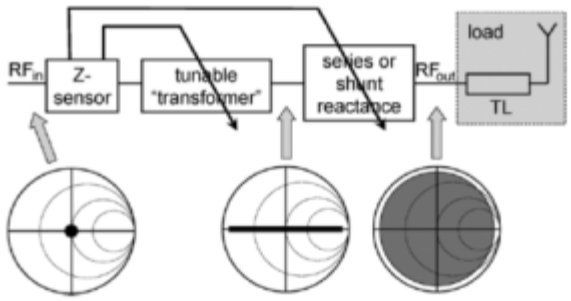


Fig. 2. Basic concept of the automatic antenna tuning system: tune out the imaginary part and tune the real part separately.

Figure 15. New AT architecture for fluctuating environment (reproduced from [9]).

In the new AT, the antenna impedance sensor makes a quasi-DC log-peak measurement of the RF signal amplitude at three points in the signal chain (V1, V2, V3 in Figure 16). Details of the log-peak detector circuit are shown in Figure 17. These data combined with the known transfer functions of the reactive elements (jX_{ext}) permit a calculation of the impedance's imaginary part. The real part of the impedance is determined from the signal level $V_{\Delta R}$ and an offset voltage that brings this level to a target value corresponding to 1:1 VSWR. The tunable transformer comprises a T-network lumped transmission line made of two adjustable inductors and an adjustable capacitor. Any appropriate device can be used to implement these elements.

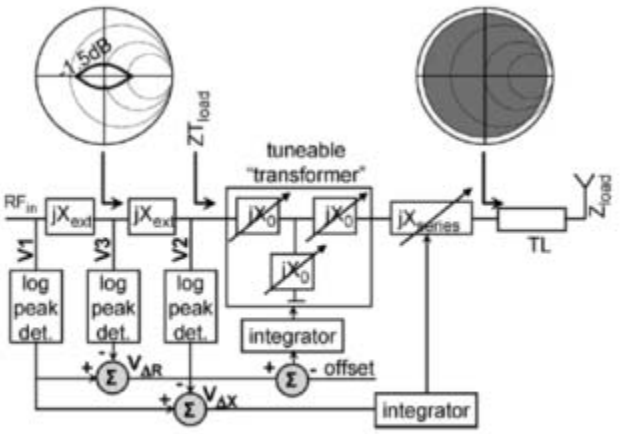


Fig. 3. Automatic antenna tuning system with low-frequency impedance detection; the target impedance (− 1.5-dB contour) is used in Section IV-A.

Figure 16. AT functional block diagram (reproduced from [9]).

Two demonstration versions of this new AT design was built and tested at 900 MHz. Measurements confirmed that they could reliably and quickly automatically tune antennas with $Z_{in} = [5-500] + j[15.8-158]\,\Omega$ (10:1 VSWR). Although the sizes were not reported, these demonstration units were intended for use in cellular handsets, so that the size is certainly consistent with use in a set-top television AT.

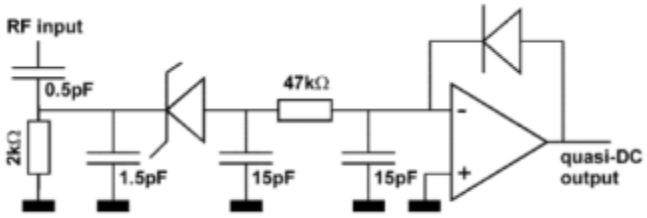


Fig. 4. Schematic diagram of the log-peak detector implementation.

Figure 17. Details of log-peak detector circuit in Figure 16 (reproduced from [9]).

2.5.1.3 On-Chip Tracking Filters

While tracking filters are not AT's, their technology likely is applicable to antenna tuners ATs, and consequently should be monitored for application in set-top devices. This section describes three different types of tracking filter.

A device developed specifically for DTV tuners is described in [10]. A new architecture is proposed that includes two separate filters, a harmonic rejection tracking filter (HRTF) and an RF tracking filter (RFTF) that are controlled by complementary digital switches. The new architecture is shown in Figure 18. The entire device is fabricated as a single integrated circuit (IC) chip. The filter stage is shown in blue located between the low noise amplifier (LNA) connected to the antenna and the DTV tuner.

Unlike narrowband RF receivers, wideband DTV tuners are prone to interference from local oscillator odd harmonics mixing with signals at the lower television channels (48-287 MHz). This problem is addressed by the HRTF that provides a high-order band-pass tracking filter with 3rd-order harmonic rejection greater than 60 dB. On the UHF television channels the RFTF tunes 287-860 MHz with narrow band response (20 MHz at -3 dB). This stage is implemented using a cascade of tunable 2nd-order bandpass tracking filters.

Computer simulation of the new architecture's performance predict 48-860 MHz operation with tunable adjustable bandwidth of 8-20 MHz, 5-15 dB N+2 channel rejection (16 MHz offset), and 3rd-order harmonic rejection of 60 dB from the HRTF. The RFTF stage provides 4.2 dB of N+2 channel rejection above 287 MHz. This device can be fabricated on-chip using 0.13 µm CMOS technology with 1.2 VDC supply voltage and 19.8 mA current draw (total of 24 mW power consumption).

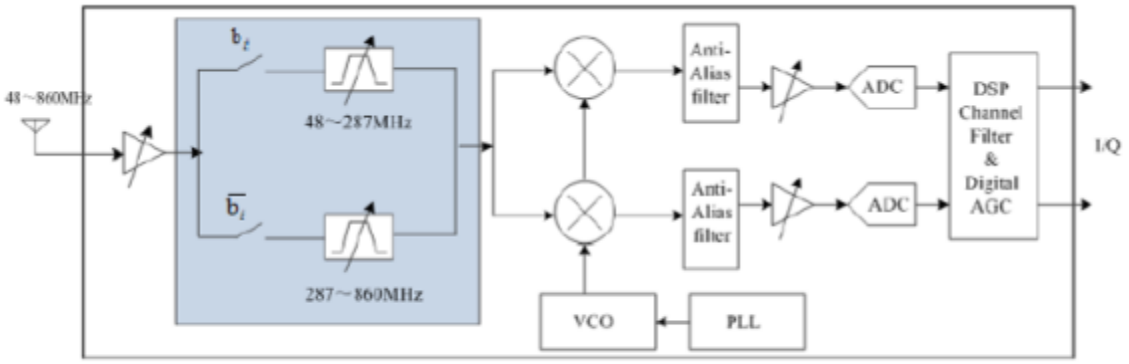


Figure 1. Direct Conversion TV tuner architecture

Figure 18. New DTV tracking filter architecture (reproduced from [10]).

Another potentially important emerging filter technology is the use of low-temperature co-fired ceramic (LTCC) technology to fabricate fully integrated multi-layer tunable filters for RF and microwave use. A typical device is shown in Figure 18. Passive

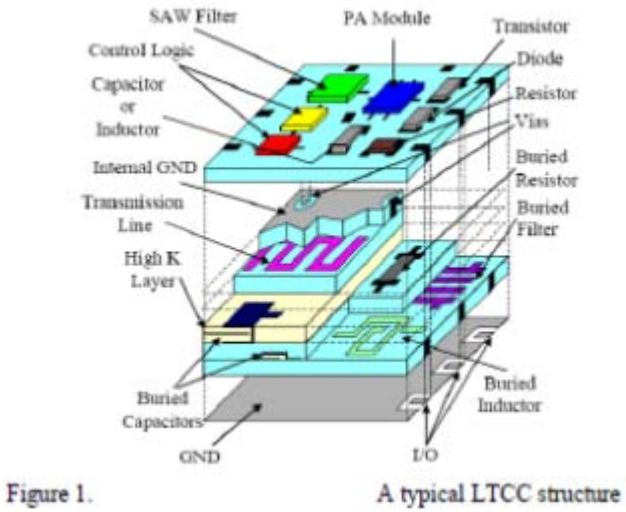


Figure 18. LTCC filter structure (reproduced from [11]).

Elements (resistors, inductors, capacitors) are integrated on the surface or embedded in a multilayer substrate. Various configurations have been demonstrated, including bandpass filters (380 MHz-2.4 GHz), 3-stage Butterworth bandpass filters (1.2 GHz), and an electronically tunable microstrip combline filter. These prototype devices point to LTCC technology's utility for RF and microwave applications. At this point the new multilayer architecture proposed in [11] that involves switching between layers for tuning is being computer-modeled, but working devices based on that approach have not been fabricated. LTCC technology may be very attractive in the television frequency range because of its potential for very high levels of integration resulting from the multilayer design.

A third example of an on-chip tracking filter is provided by [12]. A complete tunable structure was fabricated and tested. This chip occupied an area of only 2.8 mm^2 (fabricated with 0.18 µm CMOS) requiring 34-120 mA at 1.8 VDC. A photomicrograph of the chip appears in Figure 19, and its architecture in Figure 20. The device comprises cascaded RLC sections as shown in Figure 20, each containing a digitally programmable on-chip capacitor and resistor, and an off-chip fixed inductor. The resistor is adjusted with 8-bit resolution, while the capacitor uses a 10-bit control signal.

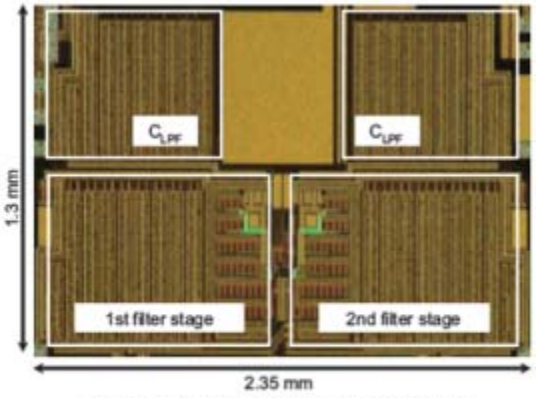


Fig. 7. Chip micrograph

Figure 19. Tunable LC-Tracking filter as fabricated (reproduced from [12]).

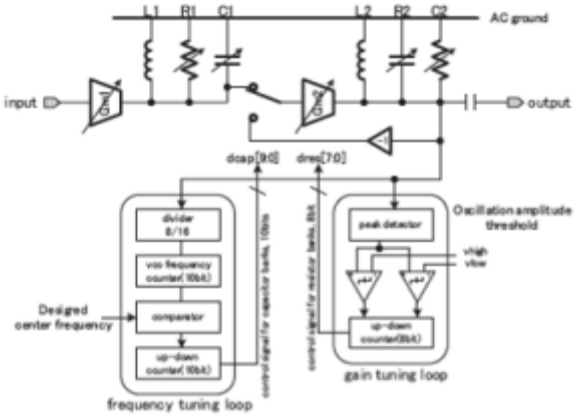


Fig. 1. RF tracking filter architecture

Figure 20. Tunable LC-Tracking filter architecture (reproduced from [12]).

Figure 21 shows the tracking filter's measured performance data from 125 MHz to 1.06 GHz. Its response in the 5.6 MHz passband is very flat with a ripple less than 0.2 dB. The noise figure in this device appears to be somewhat high (16.8-19.5 dB), but the third-order intercept points are good (~128/167 dBμV, in/out of band). Frequency selectivity is good at > 36 dB, and the power consumption quite low (<~0.2 W maximum). This example shows that very effective single-chip RF tracking filters with minimal off-chip components (in this case two inductors) can be designed and fabricated for set-top use using currently available technologies.

TABLE I
PERFORMANCE SUMMARY

Parameter		Measured value
Center frequency(f_C)		125 MHz – 1.06 GHz
Passband ripple	f_C =125MHz (BAND1)	< 0.2 dB
Signal bandwidth		5.6 MHz
Noise Figure (Input referred noise)	f_C =430MHz (BAND2)	16.8dB (17.4 dBuV)
	f_C =1.06GHz (BAND2)	19.5 dB (20.1 dBuV)
in-band IIP3		127.7 dBuV
out-of-band IIP2		167.0 dBuV
Frequency selectivity at 3rd LO freq.		> 36dB
Power consumption	f_C =380MHz (BAND2)	120mA @ 1.8V
	f_C =430MHz (BAND1)	34mA @ 1.8V
Si-area		2.8 mm2
Process		0.18um CMOS 1P6M with MIM cap

Figure 21. Measured performance of tunable LC-Tracking filter (reproduced from [12]).

2.5.1.4 Software Defined Radios

A software defined radio (SDR) is an element of a wireless communication network whose operational modes and parameters can be changed or augmented post-manufacture via software. The essential idea is that a flexible hardware layer exists whose function can be controlled and modified entirely by a computer program, as opposed to requiring hardware modifications of any kind. The SDR concept spans many radio network technologies including cellular systems, personal communications services (PCS), 3rd and 4th generation wireless (3G and 4G), mobile data, emergency services, paging, messaging, and military/government communications, and any future modifications to these existing services or entirely new ones. The FCC (Federal Communications Commission) definition is more restrictive in that it applies only to the transmitter side of an SDR. But, as a practical matter, the SDR concept applies to any wireless device whose characteristics are software-controllable, whether it be the transmitter, receiver, both, or some other element such as a modem.

SDR technology is relevant to the television set-top AT because there is a developing standard that specifically addresses the issue of “smart antennas” (SAs) in the context of SDR. This type of antenna and its associated AT may be useful for the set-top application and consequently should be monitored as an emerging technology. The high-level SDR smart antenna architecture appears in Figure 22. The hardware layer comprises M transmit antennas and N receive antennas because SDR in general supports two-way communication (in the step-top TV receive application, of course, there are no transmit antennas). Each antenna is has a separate RF/IF processing chain with the smart antenna signal processing (“waveform application”) being applied to the

baseband signal. Device drivers in the middleware layer control various programmable hardware devices, such as ASSPs (application-specific standard processors), FPGAs (field-programmable gate arrays), DSPs (digital signal processors), and GPPs (general purpose processors).

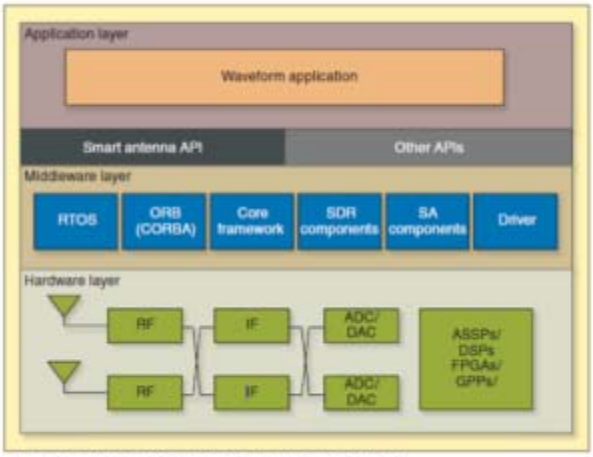


Figure 1. Example of an SDR-based smart antenna open architecture.

Figure 22. SDR smart antenna high-level architecture (reproduced from [13]).

Figure 23 illustrates a typical deployment of an SA API (application program interface). The SA control device (for example, an antenna tuner, tracking filter, MEMS-based controller, fluidic element controller, and so on) is operated by a GPP controlled by a CORBA interface (common object request broker architecture). The high-level SA algorithm controls appropriate CORBA-compatible drivers for DSP baseband processing, antenna control, and other functions such as synchronization and other SDR devices that may be controlled by the SA algorithm. The basic concept in this structure is that any hardware module involved in controlling the SA or processing its signal is controllable by the API with variable parameters.

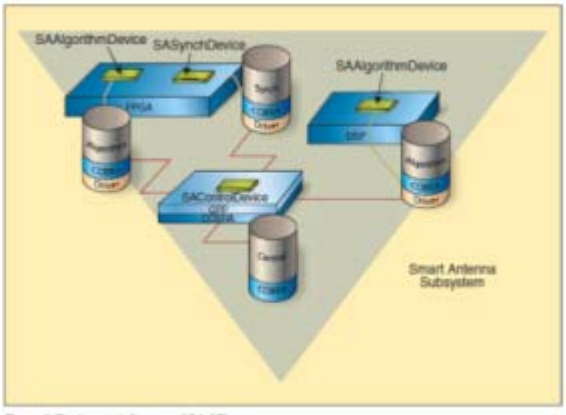


Figure 6. Deployment diagram of SA API.

Figure 23. SDR Smart Antenna API deployment (reproduced from [13]).

A typical SA hardware implementation for the SDR baseband signal processing module is shown in Figure 24. The two DSP integrated circuits (ICs), the GPP, and the FPGA ICs are labeled. This type of module is simply inserted into a backplane containing the other SDR modules to create a complete SDR. All chip-level functions are fully controllable by the SA API. As modifications are required, a simple download of the updated API is all that is necessary. No hardware modifications or swaps are necessary. The SDR approach thus provides exceptional flexibility in customizing radio performance and it well may be a very useful approach to developing effective set-top television ATs.

Figure 6: Photograph of implemented Baseband Signal Processing part of SDR smart antenna system.

Figure 24. Typical SDR smart antenna hardware implementation (reproduced from [14]).

2.5.2 References

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2.6 Physically Reconfigurable Antenna Elements

2.6.1 Summary

A *reconfigurable antenna* (RA) is an antenna whose physical and/or electrical properties can be changed in real time in order to achieve certain performance characteristics. The panoply of possible RA implementations makes RAs especially attractive for the television set-top antenna application. This section discusses some of the "typical" RA implementations, as well as some rather unusual technologies that may point the way to future developments that could be useful for set-top devices.

The essential concept underlying RAs is that some antenna property is changed "on the fly" to accomplish some objective, say tuning. An example is the tunable antenna in which the electrical length of a radiating element is changed by switching in and out reactive elements (capacitors/inductors) or remotely moving the tap on a roller-type tuning inductor. This type of reconfigurable antenna has been used for at least fifty years (see, for example, [1] p. 20.46 *et seq*.).

While an RA may be any electrical size, typically they are a substantial fraction of a wavelength, so that they are not (necessarily) electrically small antennas (ESAs). The problems associated with ESAs, narrow bandwidth, low radiation efficiency, highly reactive input impedance, are mitigated in RAs that are a substantial fraction of a wavelength in size. Nevertheless, the RA concept is applicable to ESAs as well, and it may be useful for ESA candidates for the set-top application. As a general proposition, RA technology is applicable across the entire range of antenna electrical sizes, which makes it particularly attractive for set-top devices.

Most recent examples of effective RAs are at frequencies well above the broadcast television channels, typically in the microwave region. However, because these RAs often are comparable to the operating wavelength in size, they may be readily used at lower frequencies, especially at UHF TV channel frequencies. On the lower VHF channels, RA approaches may be effective when applied to standard antennas such as bowties or spirals whose size is consistent wit set-top requirements.

Five candidate RA technologies hold promise for short or long term application to television set-top antennas:

(a) *Microelectromechanical System* (MEMS) RAs are antennas based on MEMS RF switches. These devices, which have been fabricated for well over a decade, have come to the foreground as the preferred approach for RAs. They can be used in all frequency ranges, and, in fact, perform better at lower frequencies, although current applications emphasize the high UHF and microwave ranges. Other similar switching

devices such as PIN diodes or GaAs (gallium arsenide) solid-state switches also are used to build RAs, but these technologies are older and perform less well. The latest developments in MEMS, nano-MEMS and the use of carbon nano-tubes (CNT), reduce sizes and improve performance to the point where there seems essentially no question that MEMS-based set-top RAs will be achievable in the near future.

(b) *Fluidic* RAs are a very recent development that promises simple, cost-effective, rugged reconfigurable wire antenna structures in almost any shape imaginable. Fluidic elements are fabricated from a flexible eslastomer enclosing human-hair-thin channels filled with eutectic gallium/indium (EGaln), a liquid metal. The resulting "wire" provides electrical performance similar to copper and can be stretched, bent, rolled, or twisted into almost any shape simply by applying mechanical stress. The wire returns to its original linear structure upon relieving the stress. These elements have been used to fabricate a simple center-fed dipole that is tuned by stretching. A similar approach using a multiplicity of wires or other geometries (for example, bow ties or spirals) may well be the basis of a simple, effective set-top antenna. This technology has only recently been reported, and it certainly bears watching as a candidate for set-top RAs.

(c) *Pixelated piston* RAs comprise small individually controllable and addressable "pistons" in a two-dimensional matrix. Each piston is controlled by an actuator which, in turn, is operated by an electronic controller. A conductive patch is mounted on the end of each piston, which in their quiescent position form a continuous, flat ground plane. Each piston moves back and forth above and below the ground plane to form reconfigurable antenna radiating elements or transmission lines. This technology, which is patented, has been demonstrated to provide high-gain, beam steerable, antennas operating from 500 MHz to 18 GHz with reconfiguration times less than one millisecond. Its utility for the set-top application is self-evident, but at this point in its development the cost likely is prohibitive.

(d) *Liquid crystal* (LC) RAs are fabricated using LC substrates sandwiched between two electrodes that apply a DC bias voltage to control the LC's dielectric constant by deforming its molecules. Very small unit cells are fabricated that can be formed or applied onto surfaces of various shapes. At this point in its development, LC RAs have been demonstrated only on planar surfaces, and only at extremely high frequencies (high microwave through millimeter wave ranges). An effective LC RA reflectarray antenna has been demonstrated at 77 GHz. It provided complete electronic beam steering over an angular range of 35 degrees with good gain and sidelobe performance. Whether or not this technology can be ported to low VHF television set-

top channels is not clear, but the technology bears watching because it is demonstrably effective for designing and building reconfigurable antennas.

(e) *Plasma* RAs are similar to the pixilated piston antennas, but instead of using conductive patches, they utilize ionized gas (plasma) as the conductive element. Each plasma element is turned "on" (ionized) or "off" (non-conducting) by a control voltage, so that the entire antenna aperture is electronically reconfigurable. Unlike pixilated piston RAs, the plasma elements lie in a single plane and cannot be positioned above or below the plane. This limitation, however, is minor in the context of set-top receive antennas because it is very likely that good receiving structures can be formed on a planar surface, which also simplifies the RA. The plasma RA has another potentially significant advantage over other candidate technologies. Existing plasma display technology, which is very highly developed and sophisticated, may be directly applicable to the design and manufacture of plasma RAs for the set-top application. Leveraging this existing technology may speed time to market and reduce costs substantially.

2.6.2 MEMS-based RAs

Microelectromechanical system (MEMS) radio-frequency switches have gained widespread acceptance as a standard element for implementing in RA structures. RF MEMS are routinely used to change both antenna feed networks to accomplish impedance matching and radiating element topology to control radiation pattern and efficiency. Most applications are in the microwave range, but MEMS use is not frequency-limited.

While there are other switching elements besides MEMS, notably sold-state PIN diodes and GaAs switches, MEMS offer several advantages. They are inexpensive, which is a major advantage for set-top applications, and they exhibit low insertion loss and low power consumption [2]. In recent years their availability as a commercial item has grown considerably, and they are now readily available as off-the-shelf components [3,4,5]. Demonstrated lifetimes exceed 100 billion cycles, which is consistent with consumer levels of usage over periods of several years.

Figure 1 shows a typical bi-layer curled MEMS structure. It is extremely small in size (300x1500µm). The two-layer switch element comprises materials with different thermal coefficients of expansion that pull it up away from the contact after annealing as shown in (a). The switch therefore is normally open (NO) when no control voltage is applied. Applying a DC voltage between the switch post and bottom electrode (1.5 µm thick polysilicon) creates an electrostatic field that pulls the switch element onto the contact as shown in (b). The microphotograph below the switch diagram shows its structure as actually fabricated.

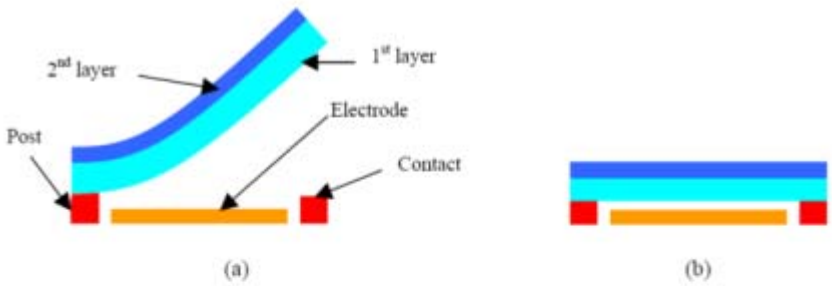


Figure 11. Operation of a bi-layer curled MEMS switch. (a) OFF state and (b) ON state.

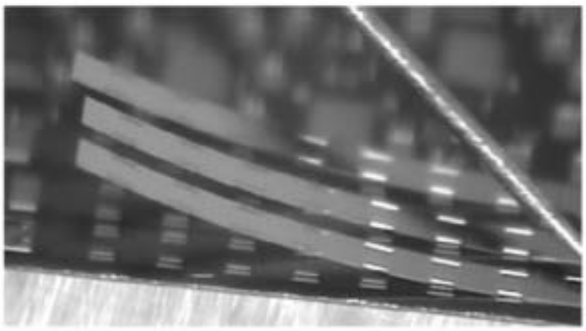


Figure 12. A photograph of the fabricated MEMS switches.

Figure 1. Typical RF MEMS switch (reproduced from [2]).

MEMS technology is improving at a rapid pace, and the latest generation of these switches are nano-scale [6]. *Nanoelectromechanical system* (NEMS) are laterally ~10-100 times smaller than a typical MEMS device, and they operate with actuation voltages below 10 volts, compared to 30-80 volts for MEMS switches. NEMS also provide faster switching times (<~1 μsec) compared to MEMS (~10-20 μsec). Figure 2 shows the new NEMS double-arm cantilevered switch schematically (a) and as fabricated (b) imaged by a scanning electron microscope.

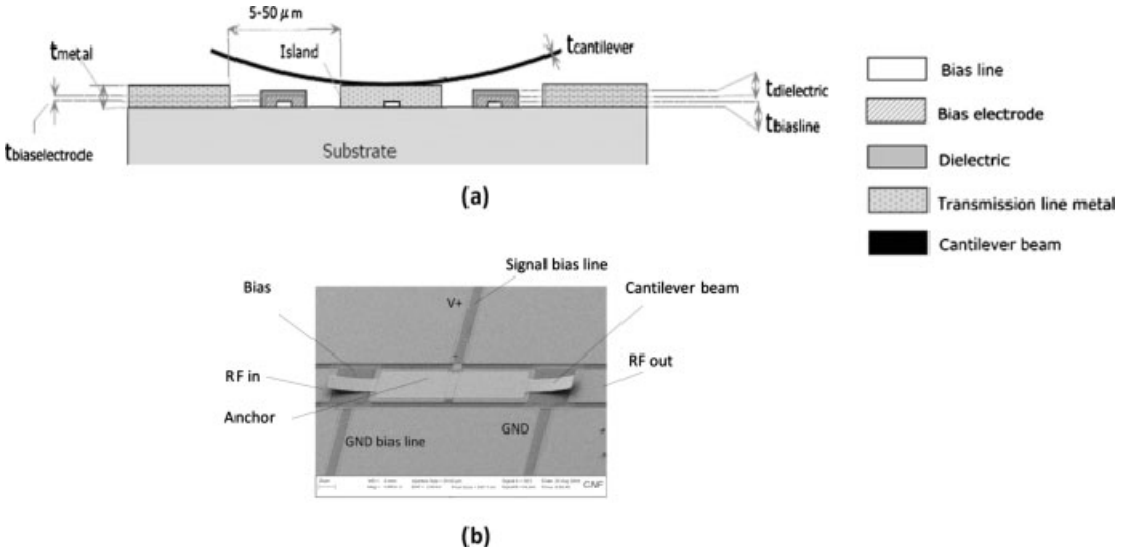


Figure 2. DC-contact NEMS switch (reproduced from [6]).

Another important recent development is the use of carbon nanotubes (CNT) to fabricate NEM switches [7]. These devices promise extreme longevity and reliability because of CNT's unusual physical and electrical properties. Carbon nanotubes are extremely strong with high electrical conductivity. In addition, their flexibility and extremely low mass suggest that switching speeds in the GHz range eventually may be attainable. Figure 3 shows two types of CNT-based NEM switches. Measured switching times were 26 ns for these devices, which while far short of the GHz range, are nonetheless faster than MEMS devices.

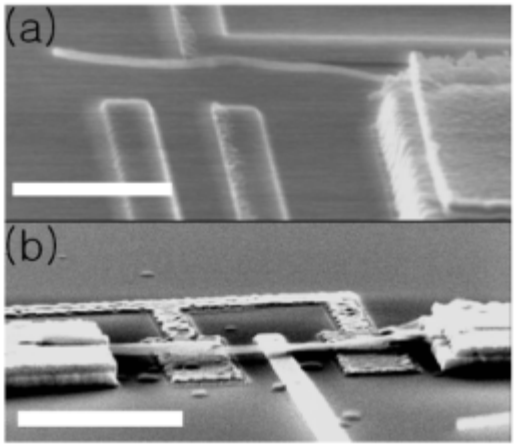


Figure 3. Two types of CNT-based NEMS switches (reproduced from [7]).

A typical MEMS-based reconfigurable array structure appears in Figure 4. This is an example of a complex RECAP (Reconfigurable Aperture) structure whose objective is wide bandwidth. Because a MEMS-based antenna can be reconfigured essentially in real-time, it can operate in multiple frequency bands. Aperture reconfiguration is achieved by switching in and out receiving and/or transmitting elements (radiating a signal, of course, is not an issue with set-top TV antennas). The structure shown in Figure 4 typically would be implemented using solid-state switches, but MEMS technology has leap-frogged that approach, so that MEMS provides better performance. In the array application, only the MEMS network departs from an otherwise standard design. The receiving/radiating element structure and its feed network are independent of the MEMS switch layer, as are the bias, FSS (frequency selective surface) and PBG (photonic band gap) layers. These last two elements may well not be required in an application like the set-top antenna because the operating frequencies are well below the microwave range where FSS and PBG structures are common.

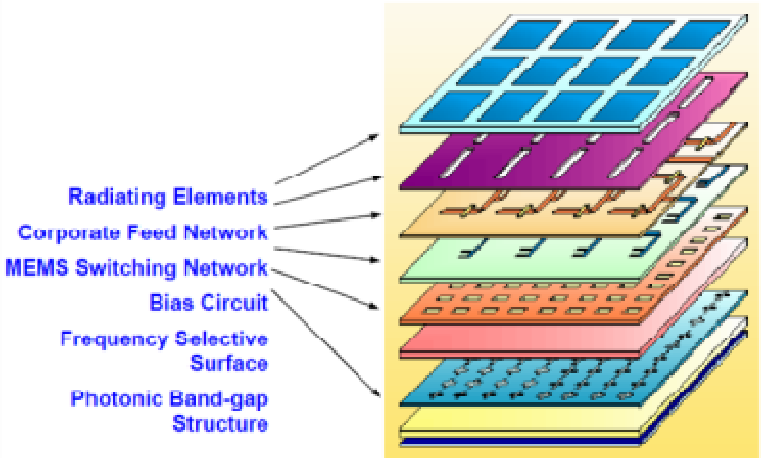


Figure 4. Reconfigurable MEMS-based antenna array (reproduced from [8]).

Another example of a state-of-the-art multiband MEMS-based antenna is reported in [9]. The design emphasizes a symmetric, repeatable topology which lends itself well to scalability, so that additional frequency bands are readily added. Scalability could be an important attribute in applying MEMS technology to the set-top environment, and this antenna demonstrates that scalability is achievable with MEMS. As configured the antenna covers four bands: 800-900 MHz; 1.7-2.5 GHz; 3.3-3.6 GHz; and 5.1-5.9 GHz. The lowest band lies just above the highest UHF television range, suggesting that this design likely could be easily modified to cover some portion of the upper TV channels. Table 2 reproduced below from [9] shows that this antenna indeed provides excellent performance in the four design bands. Its structure appears in Figure 5.

Table 2. Summary of Reconfigurable Antenna Performance

Band of interest	B1	B2	B3	B4
Frequencies covered (GHz)	0.8-0.9	1.7-2.5	3.3-3.6	5.1-5.9
Measured S11 (dB)	-13	-9	-7.7	-3.6
Measure Center Freq Gain	8.21 dB	7.96 dB	9.6 dB	8.4 dB

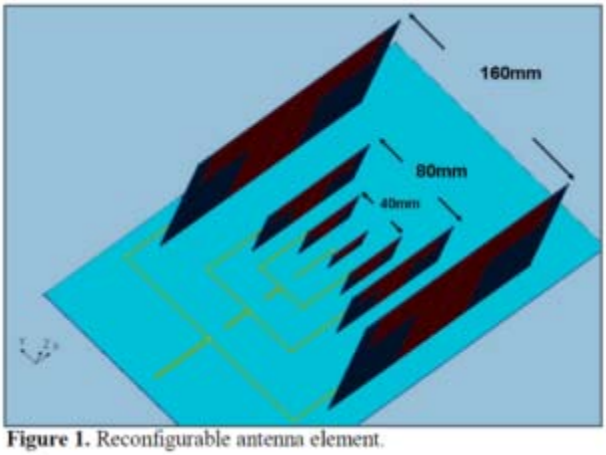


Figure 1. Reconfigurable antenna element.

Figure 5. Four band MEMS-based antenna (reproduced from [9]).

2.6.3 Fluidic RAs

An interesting new RA technology is the *fluidic* antenna. A fluid metallic alloy, eutectic gallium/indium (EGaln), comprising the metals gallium and indium, that remains liquid room temperature is injected into a microfluidic channel made from the silicone elastomer PDMS (polydimethylsiloxane). The channels are very small, about the width of a human hair, open at each end, and of any desired shape. After the channel is filled with EGaln, the alloy's surface oxidizes, creating a "skin" that holds the alloy in place while allowing it to retain its liquid properties. The alloy's mechanical properties are determined by the elastomer casing because it remains liquid.

The "wires" created by this process can be bent, stretched, rolled and twisted by applying a stress. Figure 6 shows a photograph of a twisted fluidic wire. Upon relieving the stress, the liquid conductor/elastomer reversibly returns to its original shape without hysteresis. An antenna fabricated of fluidic elements consequently is fully reconfigurable. The structure is self-healing to small cuts, and it is highly flexible and durable. Life expectancy therefore is very high, and fabrication costs reasonable because the elastomer's channels are formed using a technique known as "soft lithography" which avoids milling or etching in fabricating the fluidic antenna elements. However, the EGaln liquid metal is expensive, and at this time may be prohibitive for consumer applications.

A simple center-fed dipole was constructed from the fluidic element and tested. The antenna is shown in Figure 7. Its radiation efficiency was measured at approximately 90% over a frequency range of 1910-1990 MHz tuned by stretching the antenna. The efficiency is comparable to a copper element in this frequency range, so that the fluidic antenna is expected to perform as well as any metallic structure. In the television set-top application, this type of RA may be very attractive, especially if the price of EGaln alloy can be reduced because of volume. Of course, there are unanswered questions, for example, to what degree the elastomer casing can be stretched. The important conclusion that can be drawn from the reported research is that an entirely new fluidic technology may be close enough to maturity that it soon will be viable for reconfigurable set-top receive antennas.

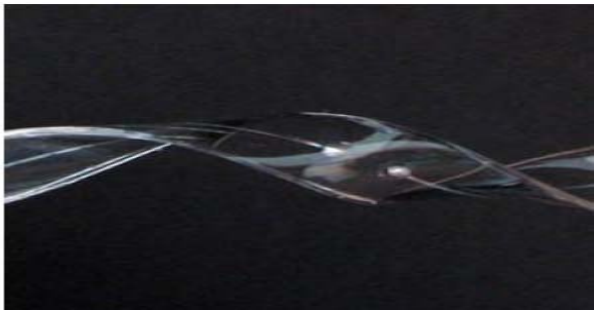


Figure 6. Twisted fluidic wire element.

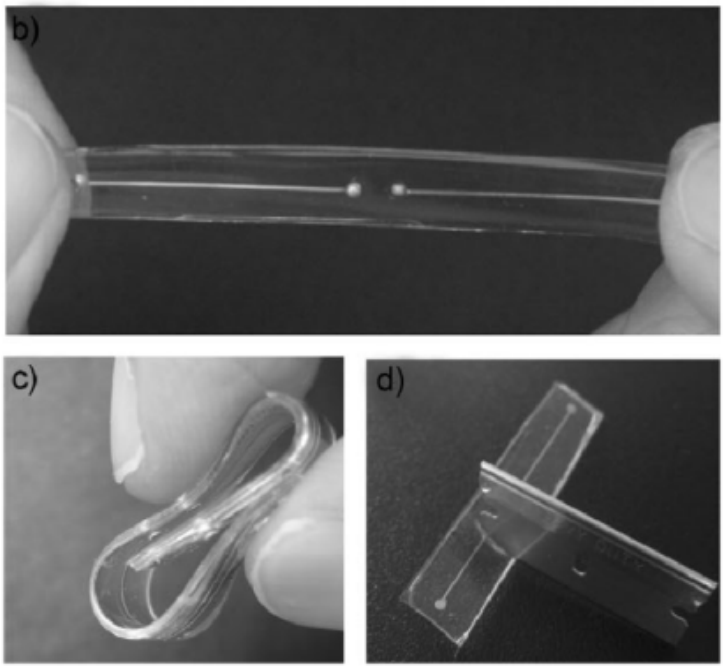


Figure 7 (a. b, c) Photographs of a prototype antenna being stretched and rolled. There is no hysteresis in the spectral properties of the antenna as it is returned to the ‘‘relaxed’’ state. d) The antenna self-heals in response to sharp cuts, such as those inflicted by a razor blade. Fluidic dipole antenna (reproduced from [10]).

2.6.4 Pixel Piston RAs

An interesting and potentially set-top feasible reconfigurable antenna comprises addressable conductive segments ("pixels") attached to the top of a movable piston in a two-dimensional array of pistons [11]. This RA appears in Figure 8. The piston includes a conductive patch (1010) sitting on top of a dielectric section (1020) that is attached to a conductive segment of the same triangular shape (1030), although other shapes are possible as well. The lower circular shaft (1040) is a dielectric material. This antenna is designed to operate at very high frequencies, 20 GHz being representative, and its dimensions are consequently commensurate. Typically the conductive pixel is about $\lambda/20$ across (0.7 mm at 221 GHz), while the height of the dielectric segment is $\lambda/10$ (1.4 mm at 21 GHz) with an overall height of about 11 mm.

The reconfigurable antenna is created by raising and lowering the pistons to form a transmission line and radiating elements in any desirable allowable pattern. If a pixel is in its lower position, it comprises an element in the array's ground plane. Otherwise, it is part of a transmission line or radiating element as shown in Figure 8 on the right. The entire structure is addressable on a pixel-by-pixel basis by a controller that operates a two-dimensional actuator for each pixel. Typical reconfiguration times are less than one millisecond, which is consistent with consumer requirements for the set-top application.

While the objective of this patented technology is electronically reconfigurable microwave patch antenna arrays, it may also be applicable to the set-top TV receive antenna if alternative element configurations are considered. For example, the pixel piston structure could be useful in creating folded element antennas, spirals, or bow-tie type elements. Each of these could be tuned to the required channel by sizing it appropriately.

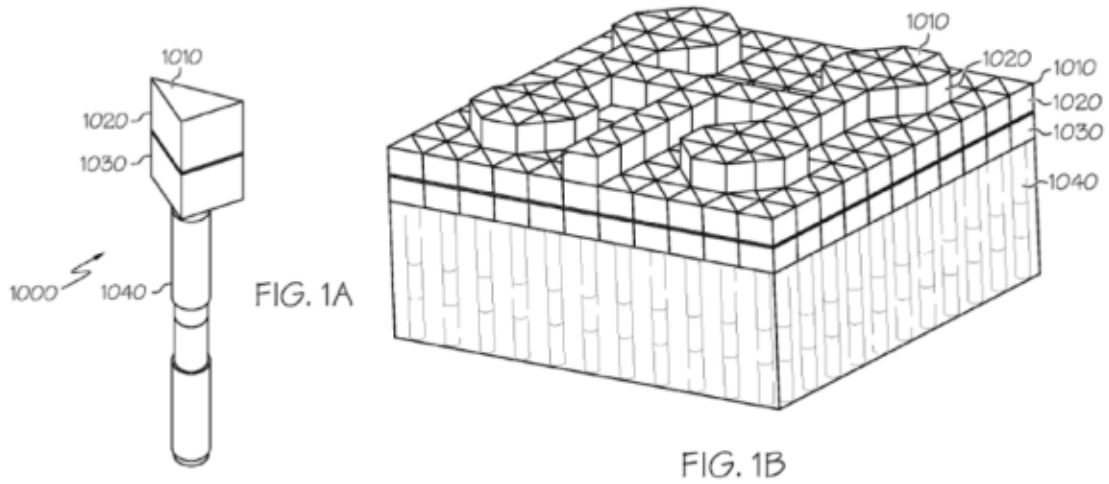


Figure 8. Pixel piston reconfigurable array (reproduced from [11]).

This technology has been implemented in commercially available antennas [12]. PARCA (Pixel-Addressable Reconfigurable Conformal Antenna) arrays are being developed for military applications requiring UWB (ultra wide-band) transmit antennas that adaptively reconfigure operating frequency, gain (beam width), and polarization while handling high power (~2 KW). A typical implementation operating from 1 to 18 GHz using approximately 100,000 ~1.65 mm diameter pixels would be 0.5 meter (20.5 in) square in size with a thickness of about 20 mm (0.8 in). Approximate weight is 12 lbs. This PARCA array would provide gain of 15 dBi and 40 dBi at 1 GHz and 18 GHz, respectively. The antenna is shown schematically in Figure 9.

Testing of microstrip transmission line and patch pixelated antennas has been done with good performance being demonstrated from 500 MHz to 18 GHz [13]. Whether or not this technology can be extended an order or magnitude lower in frequency for the set-top application with reasonable overall size is an open question. But it is clear from the current state-of-the-art that pixilated RAs constitute a technology to watch for the set-top receive antenna.

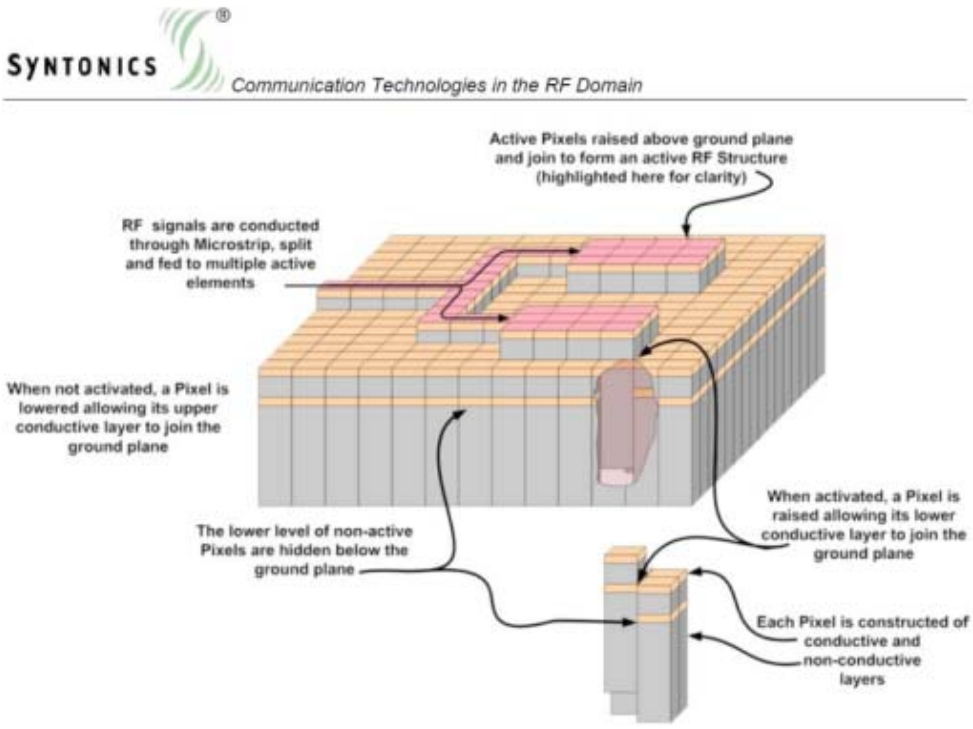


Figure 9. UWB PARCA array (reproduced from [13]).

2.6.5 Liquid Crystal Reconfigurable Antennas

Liquid crystal (LC) technology has recently been applied to RF front end components and array antennas for satellite applications, typically at quite high frequencies (> 20 GHz). Porting this technology to the low VHF range for a television set-top receive antenna simply may not be possible. Nevertheless, LC technology is worth monitoring because it may migrate to substantially lower frequencies, or it may inspire development of similar crystal-based technologies that are applicable in the VHF range.

Liquid crystals are inherently anisotropic and are so named because they simultaneously exhibit properties of liquid and crystalline materials. The class of *nematic* LCs provide the best dielectric characteristics from the microwave through millimeter wave bands, and therefore are the material of choice for very short wavelength RAs. A nematic LC molecule has a more or less rod-like shape a shown in Figure 10. Anisotropy is reflected in different perpendicular and parallel (relative to an incident electric field) dielectric constants, typical values being ~2.5 and ~2.8-3.5, respectively. Application of a bias voltage to an LC causes a deformation of the molecular alignment as shown schematically in Figure 10, the effect of which is to change the LC's dielectric properties. These changes, in turn, constitute the basis of a reconfigurable antenna element.

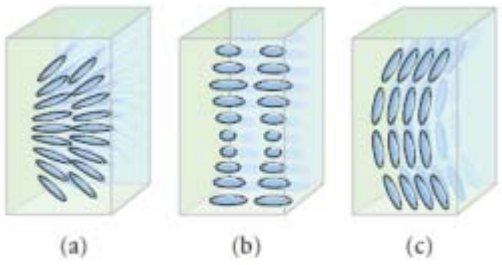
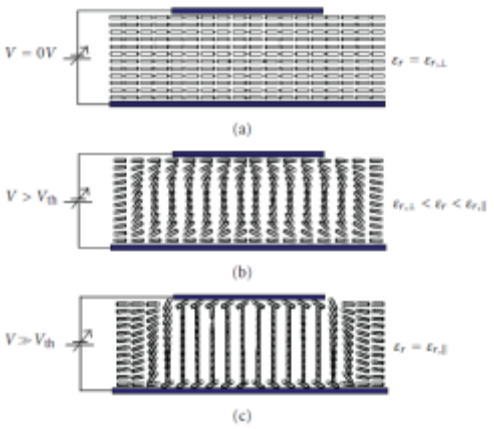


FIGURE 3: (a) The splay, (b) twist, and (c) bend deformation.

Figure 10. LC deformation modes (reproduced from [14]).

A reflectarray RA was fabricated using a dielectric substrate whose properties could be tuned using a bias voltage less than about 20 volts. The RA is a waveguide-fed planar reflecting surface capable of accurately steering the radiated beam into a wide range of directions. A unit cell comprising a single microstrip patch of dielectric sandwiched between two parallel plate electrodes is shown in cross-section in Figure 7 (left). The upper electrode patch is printed onto the dielectric substrate as is the lower electrode which acts as the antenna's ground plane. Initial LC molecule alignment is achieved using a thin (~300 nm) polyimide film on the electrode patch and ground plane. The applied external electric field created by the bias voltage reorients the LC molecules, thereby changing its dielectric properties. The figure shows the variation of dielectric constant as the bias voltage increases from zero to a value slightly exceeding the LC's threshold voltage, V_{th}, and then to a value that is much greater. As the effective dielectric constant changes, so does the capacitance per unit length of the microstrip patch, which allows the cell to be voltage-tunable as a reconfigurable antenna element.

The reflectarray antenna comprises 16x16 unit cells spaced 0.55λ. Each cell is 2.2 mm across with an LC cavity height of 50 μm. The complete antenna is shown in Figure 11, in which the individual cells are clearly visible, as are the control lines (bias lines). The antenna is fed using a right-angle bent WR-10 open-ended waveguide as shown in the figure. Figure 12 shows measured pattern data for different bias voltage levels. The main lobe is clearly steered into the desired directions of zero (blue), +25 (black), and -10 (red) degrees with good relative main lobe gain and sidelobe suppression.



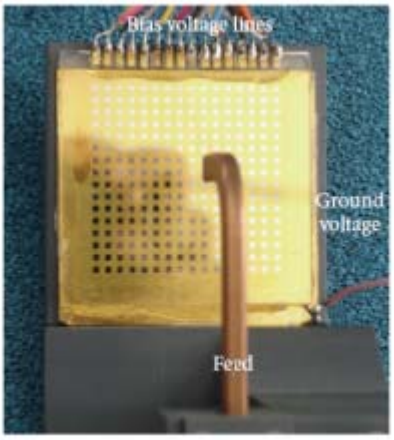


FIGURE 13: Fully assembled 77 GHz LC reflectarray with feed.

Figure 11. 77 GHz LC reconfigurable reflectarray (reproduced from [14]).

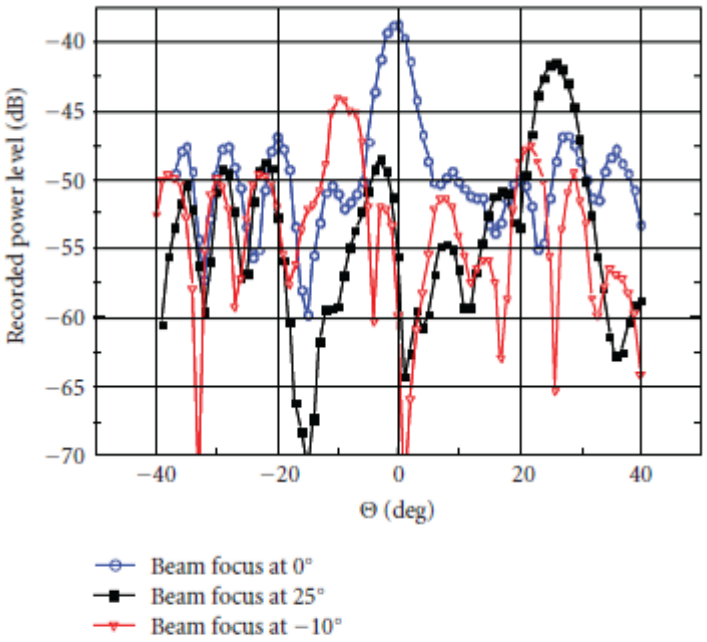


FIGURE 14: Three recorded E-plane patterns for three different voltage configurations, with the main beam pointing at different angles: −10°, 0°, and +25° (at 77.2 GHz).

Figure 12. LC reflectarray measured patterns (reproduced from [14]).

2.6.6 Plasma Reconfigurable Antennas

The plasma RA [15] is conceptually similar to the PARCA antenna in §4. It also uses electrically conductive/non-conductive surfaces to create radiating elements and associated components. But instead of conductive metallic patches, the plasma RA utilizes gas enclosures in which the gas can be made electrically conductive (plasma state) or non-conductive. The enclosures are interconnected and can be turned on (conducting) and off (non-conducting) by applying a control voltage. Figure 13 shows the basic plasma RA structure. It comprises square cells that contain four control electrodes, one on each edge. In the diagram, a cell that is turned on is shown by wavy lines between the electrodes. For example, cell 110 is "on" as shown by the wavy lines which indicate that the gas within the cell has been ionized into a plasma state. By contrast, cell 202 is "off" as indicated by the absence of wavy lines; its gaseous contents have not been ionized. Turning individual cells on and off in the manner forms the basis for a reconfigurable antenna system.

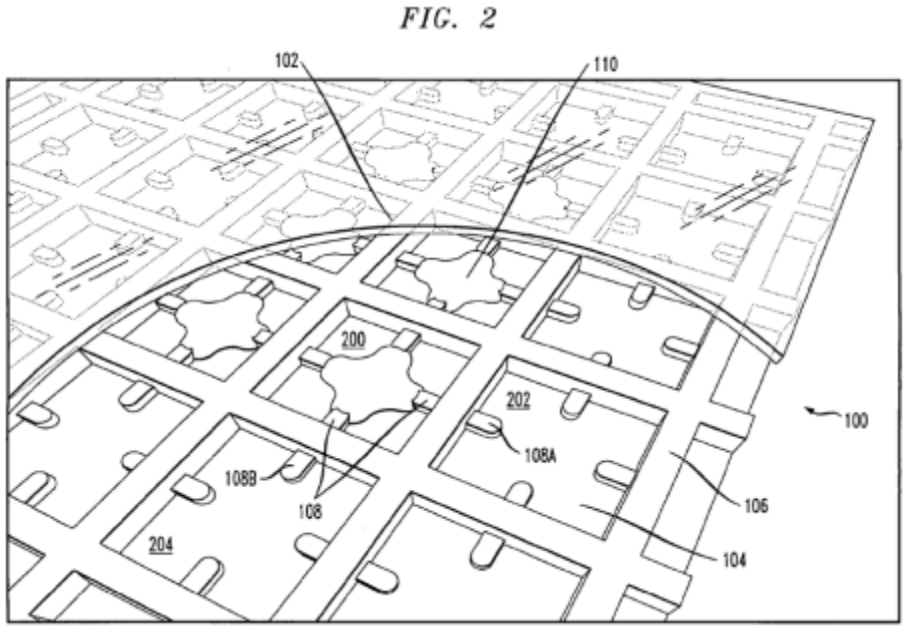


Figure 13. Plasma RA structure (reproduced from [15]).

While the reconfigurable aperture in Figure 13 can be used on a stand-alone basis, it frequently is coupled to backplane as shown in Figure 14. The backplane comprises control elements (304) that are arrayed on an insulating substrate (302) with conductive traces (312) that form the control lines. When mated with the plasma structure, the control elements turn individual gas enclosures on and off. A typical

complete plasma RA is shown in Figure 15, which adds a ground plane (404) printed on a dielectric substrate (402). The sandwich comprising the individually controllable plasma enclosures, the groundplane, and the backplane form an electronically reconfigurable antenna system.

The plasma RA may be particularly useful as a set-top antenna. One advantage is has over other approaches is that it almost certainly can benefit from the highly developed plasma screen technology used in consumer televisions. A plasma RA should be much simpler in structure and fabrication than a large plasma screen display. Applying existing display technology to manufacturing plasma RAs may be only an incremental step, thereby reducing time to market and costs.

FIG. 3

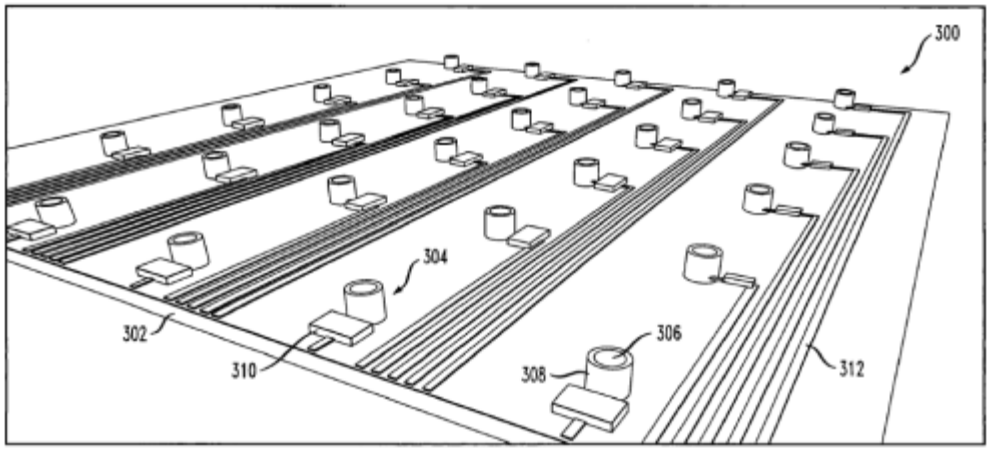


Figure 14. Plasma RA backplane (reproduced from [15]).

FIG. 4

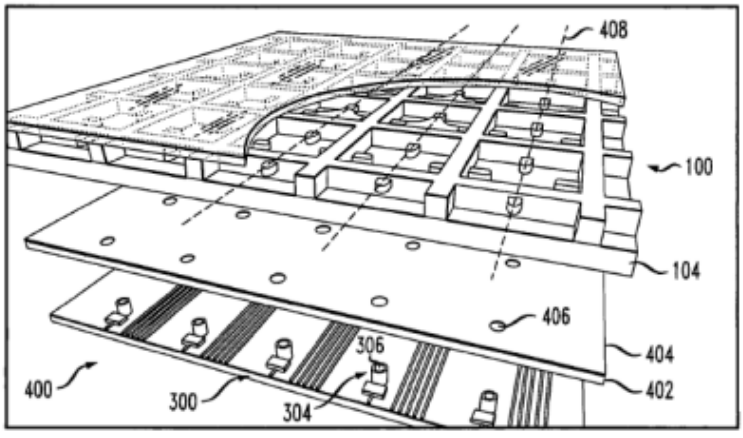


Figure 15. Complete plasma RA (reproduced from [15]).

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2.7 Use of Metamaterials in Antenna Systems

2.7.1 Summary

Traditional antennas are fabricated from electrical conductors such as copper or aluminum and insulators such as plastic or rubber. Their performance is limited by their electrical size, that is, the largest dimension measured in wavelengths. Electrically small antennas are narrow-band, inefficient, and difficult to impedance-match. The demands on a television set-top antenna are for the most part antithetical. It must be small, but it requires a very wide bandwidth to cover all VHF/UHF broadcast channels with good gain and efficiency. Traditional antennas therefore have fallen short of providing good set-top performance.

A very promising emerging technology that may soon well address this limitation is the use of *metamaterials* (MTM) in the set-top antenna. This section examines whether or not MTM can sufficiently improve the performance of electrically small antennas to make MTM-based designs attractive for set-top use. While MTM have been available since the mid-1990s and the subject of much literature, they have not gained widespread commercial acceptance until recently. A recent survey paper examines this question and suggests some explanations for the MTM's slow acceptance, much of which is not based on the technology itself.

More recent developments show that well-designed MTM-based antennas indeed can provide significantly improved performance over traditional designs, but the reported designs are in frequency bands well above the TV VHF band. Wireless communication antennas have been designed, fabricated, and measured that improve on the "Chu limit" for small antenna performance by orders of magnitude using MTM. Substantially improved bandwidth, high efficiency, and good gain are available from antennas with maximum dimensions less than one-tenth wavelength. The question is whether or not this technology can be ported to the low end of the television broadcast band; and, while the answer is not guaranteed, it appears to be in the affirmative.

The MTM-based antenna is an attractive candidate for the set-top application, and it should be vigorously studied as a candidate architecture. Because MTM typically are formed from simple wire structures, they likely can be optimized for the set-top application, which may lead to designs that are different than those used at higher frequencies that still meet set-top performance objectives.

2.7.2 What is a *metamaterial*?

The term *metamaterial* (MTM) applied to antennas does not connote a material in its commonly used sense. The materials used to fabricate an antenna broadly fall into the categories of conductors (metallic elements), insulators (non-conductors), and, occasionally, semiconductors. Materials like copper, aluminum, silver, various plastics, rubber, paper, and so on, are what come to mind as typical antenna materials; and all of these share some common features. Whether naturally occurring or man-made, they are homogeneous with electrical properties determined by their chemical structure. Like all other naturally occurring substances (at least all that are known at this time), antenna materials exhibit positive relative permittivity (dielectric constant), $\varepsilon_r \geq 1$, and positive relative permeability, $\mu_r \geq 1$. Their electrical performance in the antenna is determined by these fundamental physical parameters.

In sharp contrast, a *metamaterial* is a "composite" whose electrical properties are not determined by chemical structure, but instead by the shapes and patterns of inclusions that often are imbedded in some other material or simply arrayed in a pattern on a substrate [1]. A common inclusion is the split-ring resonator (SRR), and often combinations of SRRs and small wires are arranged to create a "unit cell" from which the metamaterial is built. Arrays of cells on some surface or within some volume constitute the actual "metamaterial." As a concrete example, Figure 1 shows a unit cell comprising a double SRR on an insulating substrate and a parallel wire. The SRR and wire are normally conducting materials such as copper or aluminum. An array of unit cells deployed on or embedded in a surface of any shape or throughout some volume becomes the *metamaterial*, which evidently is quite different from a homogeneous piece of plastic, for example, whatever its shape.

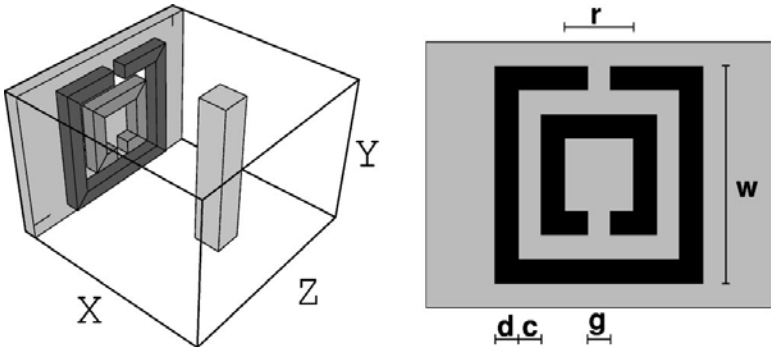


Figure 1. SRR/wire unit cell (reproduced from [1]).

What makes metamaterials attractive as antenna components is that they are *left-handed*, in distinction from all naturally occurring materials, which are *right-handed*. In right-handed materials (RHM), the electric field, $\vec{E}$, the magnetic intensity, $\vec{H}$, and

propagation (wave) vector, $\vec{k}$, are related by the "right-hand rule," *viz*, $\vec{E}$ x $\vec{H} \approx \vec{k}$. The phase and group velocities in RHM are in the same direction, parallel to the wave vector. A *left-handed material* (LHM) obeys a "left-handed rule" in which the phase and group velocities propagate in *opposite* directions. Physically this results in energy flow (Poynting vector) antiparallel to the direction of wave propagation. For LHM, both the effective permittivity and permeability are negative, $\varepsilon_r \leq 0$ and $\mu_r \leq 0$, which gives rise to LHM's unusual electromagnetic (EM) properties.

While the notion that this type of behavior can arise from as simple a unit cell as an SRR close to a wire perhaps seems counter-intuitive, it is not. There are examples of very simple wire structures exhibiting similar behavior. For example, the three-dimensional mesh of orthogonal crossed wires in Figure 2 exhibits a negative dielectric constant for frequencies $\omega < \omega_p$ where $\omega = 2\pi f$, $\omega_p = (c/\mathrm{d})[2\pi/\ln(\mathrm{d}/\mathrm{w})]^{1/2}$, and c is the velocity of light [2]. Thus, properly arrayed unit cell structures comprising conducting and insulating elements simply exhibit a similar behavior that is determined by the specific details of how the unit cell is structured.

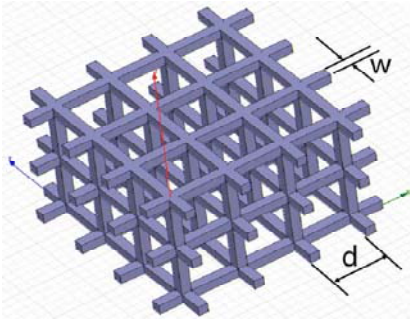


Figure 2. Crossed-wires metamaterial (reproduced from [2]).

Without doubt, the most important consequence of having $\varepsilon_r \leq 0$ and $\mu_r \leq 0$ is that the metamaterial's refractive index is negative, $n < 0$, which leads to some very non-intuitive EM effects. Besides energy flow opposite to the propagation direction, $n < 0$ results in *negative refraction* as shown in Figure 3. In the figure, air lies above the interface with either LHM or RHM. For both types of lower materials, a reflection takes place at the interface that follows the usual law of reflection (incidence and reflection angles being equal, rays 1 and 2). If the lower medium is RHM (normal material), the propagating ray (#4) is bent towards the normal. This is the usual refraction that takes place with all normal materials, and, for example, bends the image of a pencil in water towards the vertical. But, if the lower medium is LHM, just the opposite occurs. The propagating ray (#3) is bent *away* from the vertical, not towards it.

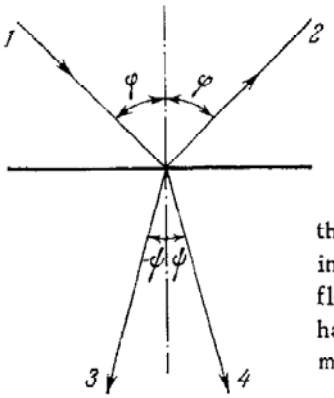


FIG. 3. Passage of a ray through the boundary between two media. 1 – incident ray; 2 – reflected ray; 3 – reflected ray if the second medium is left-handed; 4 – refracted ray if the second medium is right-handed.

Figure 3. LHM and RHM rays at air-metamaterial interface (reproduced from [3]).

2.7.3 Slow Acceptance of LHM Cellular Antennas

While there is no question that metamaterials offer intriguing possibilities for novel antennas, their use has not become widespread. Some possible explanations for this situation are offered in [4], which specifically addresses application to mobile-phone antennas. Because the cellular band (825-894 MHz, $\lambda = 0.36 - 0.335m$) lies just above the broadcast television band (54-806 MHz, $\lambda = 5.55 - 0.37m$), the issues raised in [4] are largely applicable to TV antennas.

An important consideration in whether or not LHM are useful is the antenna's *electrical size*, that is, its size measured in wavelength, λ. The wavelength at the highest TV channel 69 is the same order as the cell phone wavelengths, whereas TV channel 2 has a wavelength about 15 times larger than the longest cellular wavelength. This observation raises the question of whether or not general statements or conclusions made about antennas for cellular frequencies hold up for low-band TV antennas. The answer is a clear "maybe."

A typical cell phone antenna with a largest dimension of 0.1m has an electrical size of 0.28λ at 825 MHz. A channel 2 TV antenna has the same electrical size if its largest dimension is 1.55m, which is too large for a set-top antenna. A more reasonable set-top dimension might be 0.5m, which is 0.09λ at 54 MHz. So the question is whether or not observations about a 0.28λ antenna can be applied to one that is about three times smaller. There is no definitive answer, but the electrical sizes are close enough that some attributes of metamaterials applied to cellular antennas can be ascribed to TV antennas as well. Thus, some of the reasons in [4] why LHM are not widely used in cellular antennas may apply to TV antennas as well. Among those limitations are:

1. *Bandwidth.* Metamaterials tend to exhibit strong resonances accompanied by substantial dispersion and losses. The result is a high-*Q*, narrow-band structure that makes it very difficult to cover the entire range of VHF and UHF television frequencies. An antenna covering all cellular frequencies has a fractional bandwidth (BW) of 8% (relative to band center). If a set-top TV antenna contains two antennas, one low-band (channels 2-13, 54-216 MHz) and the other high-band (channels 14-69, 470-806 MHz), they must have fractional bandwidths of 120% and 53%, respectively, to cover all channels. If achieving 8% BW in a LHM cellular antenna is difficult, then meeting the requirements of a set-top antenna is even more difficult.

2. *Other Factors.* Whether or not an antenna is "good" for a specific application depends on many factors, many of which are not electrical in nature. Antennas that perform well electrically (gain, bandwidth, efficiency) may be too costly, too large, too heavy, too difficult to manufacture, and so on. Reports of successful LHM-based antennas usually do not address these factors, and as a result it may be difficult to analyze the benefits and risks associated with that type of antenna. This consideration has slowed the adoption of LHM antennas for mobile phones and likely applies to TV set-top antennas as well.

3. *Marketing*. Like any consumer product or component thereof, LHM antennas must be properly marketed to be effectively used. Many types of man-made materials are used in antennas, especially in the microwave range, and the resulting designs are considered "traditional." LHM-based designs often resemble traditional ones, so that they share many common features. Yet LHM antennas are frequently described as being fundamentally "different," which leads to some measure of reluctance in adopting them. A better approach would be to characterize LHM antennas as *alternatives* to traditional designs rather than a fundamental departure.

2.7.4 Examples of LHM Antennas

Even though the speculation in [4] provides possible reasons why MTM-based antennas are not more widely used, recent developments show rather convincingly that technical objections are not valid reasons. Quite to the contrary, state-of-the-art MTM antenna designs support the view that they well may be able to meet set-top TV performance requirements. This section consequently examines some typical designs in the context of porting them to the low VHF television channels.

(a) *Compact Broadband Monopole.* An example of the state-of-the-art in small broadband LHM antennas is shown in Figure 4(a), which is reproduced from [5]. The antenna comprises a PCB (printed circuit board) monopole loaded with a left-handed

negative-refractive-index transmission-line (NRI-TL) metamaterial. The loading reduces the electrical size of the antenna while providing large bandwidth.

This particular antenna was designed for dual-band use in the WiMax and WiFi bands [3.3-3.8 GHz, ($\lambda = 0.09\text{m} - 0.08\text{m}$) and 5.15-5.85 GHz ($\lambda = 0.06\text{m} - 0.05\text{m}$), respectively]. Its physical size of 0.03m [W_g in Figure 4(a)] corresponds to 0.3λ at the lowest frequency, which is comparable to the electrical size of a cellular antenna and about three times larger than a typical 0.5m set-top antenna at channel 2. Adding NRI-TL loading increased BW over the unloaded structure by approximately 1.5 GHz without increasing its size. The LHM monopole's BW (return loss below -10 dB) is 3.84 GHz from 3.15-6.99 GHz (fractional BW of 76%). The computed radiation efficiency is close to 90% across both the WiMax and WiFi bands. Input impedance is plotted in Figure 4(b), and it shows moderate values across a very wide frequency range, from about 5GHz to 10 GHz, so that good impedance matching should be readily achievable without requiring a balun.

The compact LHM monopole provides a good example of enhancements directly attributable to NRI-TL loading of a simple structure: excellent impedance bandwidth and high efficiency in an electrically small structure. Although this particular antenna would not meet a set-top TV antenna's requirement of 120% BW (see §3), it certainly suggests that with further improvements that objective might be reached.

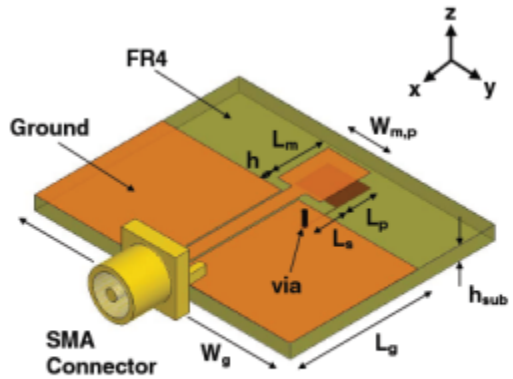


Fig. 1. NRI-TL metamaterial monopole antenna. All dimensions are in mm: $L_m = 6$, $W_m = 5$, $L_g = 15$, $W_g = 30$, h = 1, $L_p = 4$, $W_p = 5$, $L_s = 3.45$, $W_s = 0.1$, $h_{sub} = 1.59$, $S_{cpw} = 0.2$, $W_{cpw} = 1.55$, via diam. = 0.5.

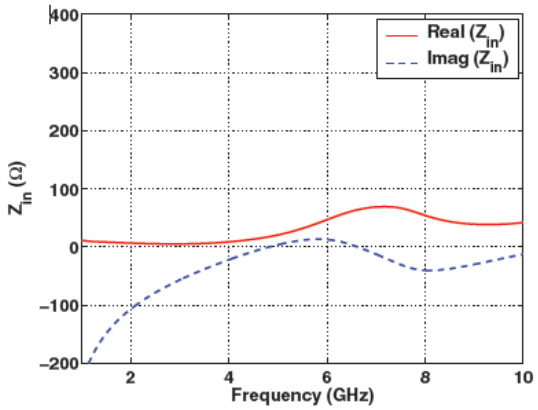


Fig. 4. Simulated input impedance characteristics of the unloaded monopole antenna from Ansoft HFSS.

(a) (b)

Figure 4. Broadband LHM monopole: Structure (a); Input impedance (b) (reproduced from [5]).

(b) *Efficient Electrically Small Antenna (EESA).* An MTM-based Efficient Electrical Small Antenna (ESSA) [6] is shown in Figure 5. It provides another example of how LHM antennas can achieve performance that is not possible with traditional designs, and suggests that porting MTM designs to TV set-top antennas is a viable approach to achieving improved performance.

The ESSA comprises a monopole surrounded by an SNG (single negative) MTM structure mounted on a plane substrate. It achieves a voltage standing wave (VSWR) bandwidth ≤ 2 of 470 MHz (1.75-2.22 GHz) [fractional VSWR bandwidth of 24%] in a package with an electrical size of only 0.07λ. The addition of LHM in this case resulted in a much broader bandwidth than could achieved otherwise because the effect MTM's effect is to lower the system *Q*-factor from the theoretical "Chu limit" of 16.72 for traditional materials to 2.23. Significantly, the increased bandwidth is not accomplished at the expense of radiation efficiency or gain, which often must be sacrificed in traditional designs. The ESSA's measured maximum gain and efficiency were 2.44 dBi and 83.61%, respectively.

The radiation patterns generally exhibited broad lobes with fairly shallow nulls ($<\approx -25\text{dB}$, Figure 6), which lend themselves well to omnidirectional reception as required in a set-top antenna. Because the ESSA's electrical size is comparable to that required for a set-top antenna (see §3), it may be possible to "stack" several ESSA's covering different band segments in a fairly small set-top package. This particular ESSA design suggests that MTM-based antennas may well meet TV set-top reception requirements.

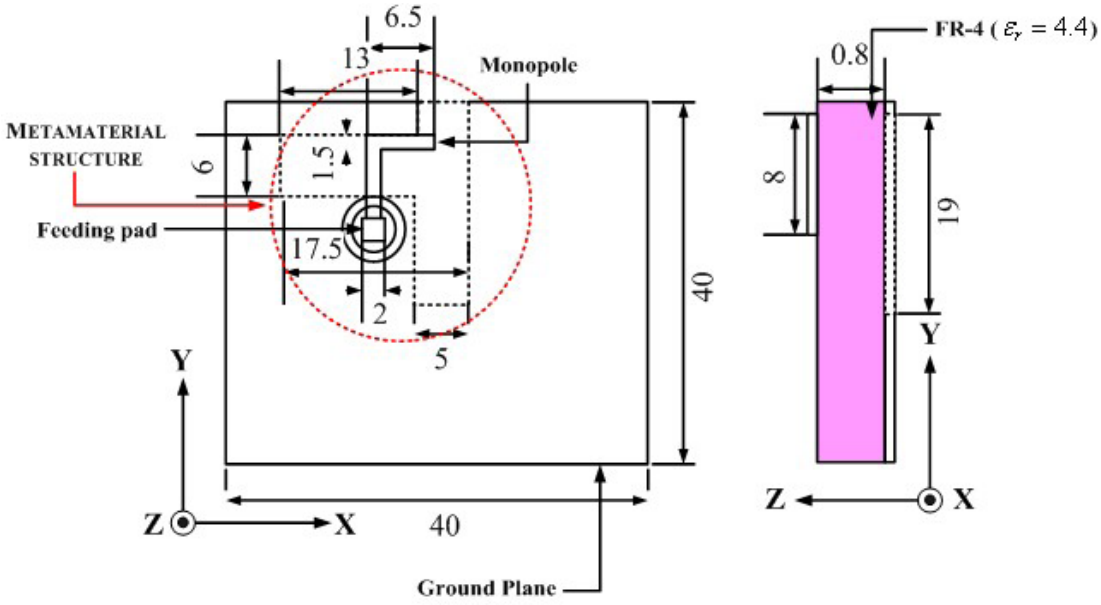


Figure 5. MTM ESSA geometry (reproduced from [6])

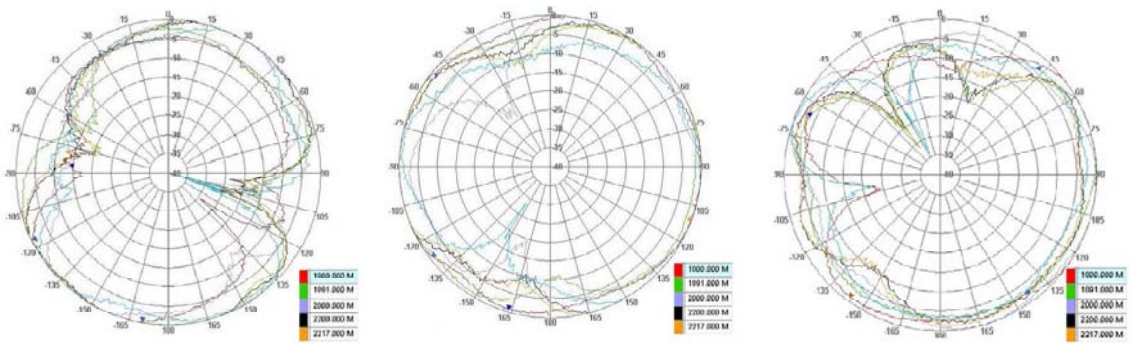


Figure 6. ESSA measured radiation patterns (reproduced from [6]).

(c) *LHM Topologies.* There are some general approaches to MTM-based antennas that appear to be useful regardless of the specific type of radiating element [7], and these likely should be the focus of future efforts to apply LHM to television set-top antennas. Various types of materials are available as shown in Figure 7. "Normal" materials have positive permittivity and permeability, and are referred to as Double Positive (DPS). "Epsilon negative" (ENG) materials can occur naturally in unusual states of matter (for example, plasmas), while *homogeneous* "mu negative" (MNG) materials can be man-made (for example, ferrite). LHM materials, on the other hand, do not occur in Nature and are "Double Negative" (DNG) because both permittivity and permeability are negative. While composite MTMs usually are DNG, ENG and MNG types also can be made and sometimes are desirable in antenna applications.

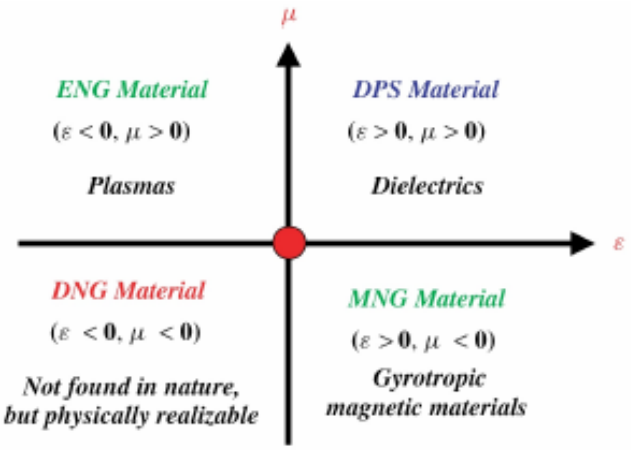
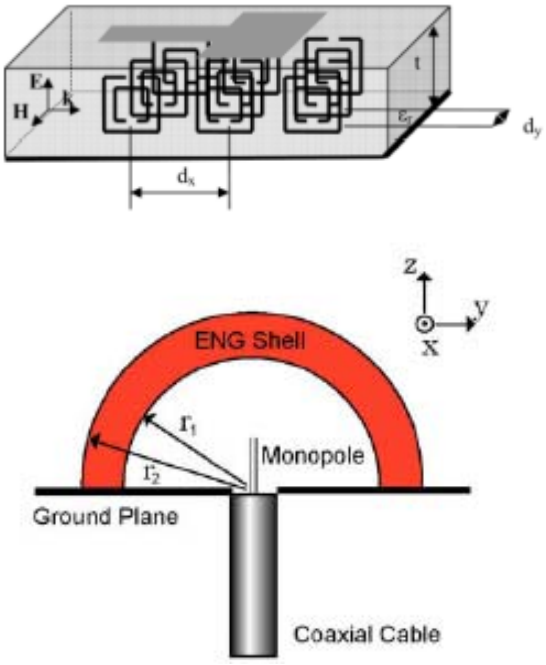


Figure 7. Material classification (reproduced from [7]).

Figure 8 shows two LHM configurations that should be useful in a wide range of applications and, in particular, TV set-top antennas. In Figure 8(a) a PCB antenna is placed on an MTM substrate in order to decrease radiating element size while in creasing radiation efficiency and bandwidth. In this example, the substrate is an MNG

array of square SRRs imbedded in a homogeneous dielectric medium. Figure 8(b) shows an ENG shell surrounding a radiator fed against a PEC (perfectly electrically conducting) ground plane. The shell can be very thin ($<\approx 0.01\lambda$) so that it has little effect on attenuating signal level, yet it substantially reduces antenna input reactance over a fairly wide range of frequencies. While the "Chu *Q*-factor" is the recognized theoretical limit for antennas made from normal materials, the ENG shell can lower *Q* by a factor of 1,583 while providing a good match to 50 Ω and efficiencies approaching 98%, at least over some range of frequencies. These examples illustrate LHM's potential benefits for TV set-top antennas, and they should be vigorously pursued for the next generation of set-top antennas.



(a) (b)

Figure 8. Generally useful LHM configurations (reproduced from [7]).

2.7.5 LHM Antenna Patents

An online search at Patent Storm (http://www.patentstorm.us/) using the keyword 'metamaterial' returns 521 U.S. issued patents and applications, many of which are related to antennas or optics. The application of MTM to antennas is a very active technology area, both in the scientific and engineering literature and in the realm of patented proprietary technology. While a thorough review of the patent literature is beyond the scope of this report, it is instructive to examine a specific development as representative of the state-of-the-art.

Figure 9 shows a typical implementation disclosed in U.S. patent #7,592,957 B2, which issued recently (Sept. 22, 2009) and consequently representative of the state-of-the-art. The figure illustrates one implementation of many described in the patent (forty-nine drawing sheets, many with multiple figures). This patent discloses composite right- and left-handed (CRLH) MTM structures as the basis of electrically small antennas and antenna arrays for use in wideband multi-band communications. One design, for example, provides 5.6 dBi gain with a 2.35-4.42 GHz BW (S_{11}<-10 dB). This level of performance is representative of what can be achieved today for wireless communication applications, and it is quite good compared to the performance of traditional antennas. It bodes well that LHM-based antenna designs likely can be useful for the TV set-top application.

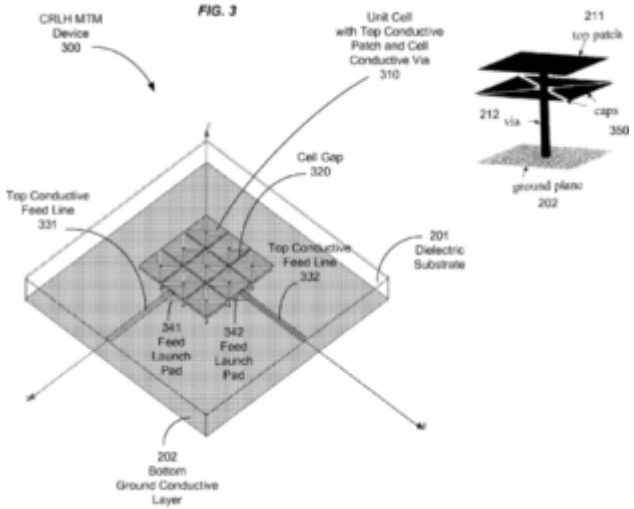


Figure 9. Typical CRLH MTM 2-D array (reproduced from U.S. patent #7592957)

2.7.6 Theory of LHM

A *metamaterial* (MTM) or *left-handed material* (LHM) has simultaneously negative values of relative permittivity (ε_r) and permeability (μ_r). There are no known naturally occurring substances with this attribute, but there are artificial media exhibiting left-handedness. MTM are useful antenna components because unusual electromagnetic effects occur when ε_r and μ_r are both negative. For example, the direction of power flow, wave refraction, Doppler shift, and the direction of Cherenkov radiation are all dramatically different than they are in "normal" media. These and perhaps other unusual electromagnetic properties of MTM may be useful in antenna system design, particularly in the microwave region.

Artificial MTM can be manufactured from arrays of conducting elements, such as small wire segments, split ring resonators, or other similar structures. These discrete elements may mimic continuous media, because, surprisingly, their left-handedness can extend to frequencies where the wavelength is much larger than either the element size or spacing. MTM slabs may be useful as coatings on metallic antenna parts. If the metamaterial can be "worked" appropriately, then major antenna components might be fabricated from it directly. The utility of these materials is an open question, because they are not well understood either theoretically or experimentally. In fact, serious investigation of these materials has occurred only in about the past decade. What follows is a discussion of how metamaterials behave and some of the issues in using them as antenna components.

Maxwell's Equations

All electromagnetic (EM) phenomena, including the EM behavior of metamaterials, are described by Maxwell's equations. Maxwell's equations are completely general, and apply to EM phenomena in all media at all frequencies, including electrostatics ("zero" frequency). MTM behavior must be consistent with Maxwell's equations in every limiting case for which theoretical results are available, whether electrostatic or time-varying. Limiting cases may be useful in understanding the counter-intuitive behavior of metamaterials.

Maxwell's equations are:

$$\nabla \cdot \mathbf{E} = \frac{\rho_t}{\varepsilon_r \varepsilon_0} \tag{1}$$

$$\nabla \cdot \mathbf{B} = 0 \tag{2}$$

$$\nabla \times \mathbf{E} + \frac{\partial \mathbf{B}}{\partial t} = 0 \tag{3}$$

$$\nabla \times \mathbf{B} - \varepsilon_r \varepsilon_0 \mu_r \mu_0 \frac{\partial \mathbf{E}}{\partial t} = \mu_r \mu_0 \mathbf{J}_m \tag{4}$$

in which: $\mathbf{E}$ is the electric field (volts/meter).

$\rho_t = \rho_f + \rho_b$ is the total charge density (coulombs/cubic meter).

ρ_f is the free charge density, and ρ_b is the bound charge density.

$\vec{B}$ is the magnetic induction (teslas or webers/square meter).

$\mathbf{J}_m = \mathbf{J}_f + \partial\mathbf{P}/\partial t + \nabla \times \mathbf{M}$ is the total current density (amperes/square

meter), consisting of $\mathbf{J}_f$, the current density of free charges; $\partial\mathbf{P}/\partial t$, the polarization current density; and $\nabla\times\mathbf{M}$, the magnetization current density.

$\mathbf{M}$ is the magnetization (amperes/meter).

ε_0 is the permittivity of free space (farads/meter).

ε_r is the relative permittivity, or dielectric constant.

μ_0 is the permeability of free space, which is arbitrarily set equal to $4\pi\times10^{-7}$ teslas meter/ampere (henrys/meter).

μ_r is the relative permeability.

In materials where Ohm's law applies, the free current density is related to the electric field by $\mathbf{J}_f=\sigma\mathbf{E}$, where σ is the conductivity (Siemens/meter).

Physically, the two fundamental vector fields are $\mathbf{E}$ and $\mathbf{B}$, because it is these fields that produce forces on electric charge according to the Lorentz force law $\mathbf{F}=q(\mathbf{E}+\mathbf{v}\times\mathbf{B})$. It is customary, however, to write Maxwell's equations in terms of the fields $\mathbf{E}$ and $\mathbf{H}$, where $\mathbf{H}$, the "magnetic field" or "magnetic intensity" (amperes/meter), is proportional to $\mathbf{B}$. The primary reason for doing this is that the quantity $\mathbf{E}\times\mathbf{H}$ is the energy flux density, which is of paramount importance is determining a system's electromagnetic behavior.

Media Parameters

It is necessary to specify the electrical and magnetic properties of the medium supporting the electromagnetic field in order to solve the field equations, subject to appropriate boundary conditions. Without this information, Maxwell's equations do not form a determinate system.

The EM behavior of any medium is determined by its material properties, the electric permittivity and magnetic permeability. There are other material parameters, but they are secondary in nature. For example, in linear media the electric susceptibility is the proportionality factor between the applied electric field and induced polarization. Similarly, the magnetic susceptibility relates magnetization to the applied magnetic

field. Each of these parameters is included indirectly in the relative permittivity and permeability, respectively, and therefore they have no independent significance.

Various classes of media exit. The simplest is *free space*, in which the relative permittivity and permeability are both unity. The EM properties of *isotropic* media are independent of the direction of the EM fields. Media in which the electrical properties are the same everywhere are *spatially homogeneous*. In some media, the EM properties vary in time, and such media are described as *temporally inhomogeneous*. The most complex media are *anisotropic*. The EM properties are direction-dependent, and may also vary spatially or temporally. The permittivity and permeability of anisotropic media are described by tensors instead of real or, in the case of lossy media, complex scalars.

Wave Velocity and Refractive Index

Time-varying fields in the form of propagating electromagnetic waves follow directly from Maxwell's equations. The equations are combined to obtain a "wave equation" whose solutions are coupled electric and magnetic field that move through the medium carrying energy as they do.

In a lossless medium, ε_r and μ_r are real numbers. A monochromatic EM wave's phase velocity, the velocity at which the wavefront (surface of constant phase) propagates, is completely determined by these two material parameters. The propagation velocity is

$$v = \frac{1}{\sqrt{\varepsilon_r \varepsilon_0 \mu_r \mu_0}} = \frac{c}{\sqrt{\varepsilon_r \mu_r}}, \tag{5}$$

where c is the velocity of light in free space, or simply "speed of light" (very nearly 3×10^8 meters/second).

In lossless "normal" media, ε_r and μ_r are positive real numbers greater than 1 (this is not true in a metamaterial). Waves in normal media always propagate slower than the speed of light. The velocity ratio is the medium's *refractive index*, given by

$$n = \frac{c}{v} = \sqrt{\varepsilon_r \mu_r}\ . \tag{6}$$

When n is real, EM waves can propagate in the medium. If n is purely imaginary at some frequency, then no wave mode (field configuration) can propagate at that frequency.

In lossless anisotropic or inhomogeneous media, ε_r and μ_r vary with propagation direction or position, and are represented mathematically as matrices of positive real numbers. In lossy materials ε_r and μ_r are complex numbers with non-zero imaginary parts that result in energy dissipation as the wave propagates through the medium. In such media the refractive index varies with direction or position or both.

n^2 may be negative (n is imaginary) in some naturally occurring media. At frequencies below its plasma frequency, for example, a gaseous plasma has a negative dielectric constant but positive permeability, resulting in $n^2 < 0$. This situation occurs in the earth's ionosphere, and provides an example of unusual wave propagation resulting from the negative permittivity. Instead of being refracted, as it would be if n^2 were positive, below the plasma frequency an incident radio wave is instead reflected. This is the mechanism that supports long distance high-frequency (HF) radio communication.

In normal media, both ε_r and μ_r are frequency-dependent, often strongly so. For example, the dielectric constant of water decreases from 81 at DC and very low frequencies, to 78 at 1 Mhz, 34 at 10 Ghz, and about 2 at optical frequencies. The refractive index varies correspondingly from 9 at DC to 1.4 in the optical region. In magnetic materials (water is not one), the frequency variation of relative permeability can be several orders of magnitude.

The effective ε_r and μ_r of metamaterials are also are frequency-dependent (in fact they must be), and they change sign just as the dielectric constant does in a plasma. As a result, in certain frequency ranges the metamaterial's novel EM properties disappear, and there is a strong analogy between the behavior of a metamaterial and naturally occurring plasma. The variation of permittivity and permeability with frequency may be very important in practical antenna applications, particularly in view of the wide spectrum used by modern communication systems. Metamaterials may be quite useful in the microwave range, but less so at lower frequencies.

Right and Left Handed Media and Power Flow

Another important parameter describing the propagation of an EM wave is its *wave vector,* $\mathbf{k}$, whose magnitude is the *wavenumber*. $\mathbf{k}$ is in the direction of propagation and perpendicular to the wavefront. In normal media, not in a metamaterial, it also coincides with the direction of energy flow.

The wavefront consists of electric and magnetic fields which are orthogonal to each other, and to the direction of propagation. The triad of vectors $\mathbf{E}$, $\mathbf{H}$, and $\mathbf{k}$ constitute a mutually orthogonal set related by the "right hand rule". When the vector

$\mathbf{E}$ is rotated into the direction of $\mathbf{H}$ through the included angle ϕ in the plane containing $\mathbf{E}$ and $\mathbf{H}$, curling one's fingers in the direction of rotation with the thumb extended perpendicular to the fingers determines the direction of the wave vector $\mathbf{k}$ (hence, right "hand" rule). The *handedness* of a medium is determined by the vector triplet $\mathbf{E}$, $\mathbf{H}$, and $\mathbf{k}$. The right hand rule relationship holds in normal media, which are consequently described as *right handed*. A metamaterial, by contrast, is *left handed* as discussed below.

The direction of energy flow in the wavefront is $\mathbf{E}\times\mathbf{H}$, which for a plane wave in normal media corresponds to the direction of the wave vector $\mathbf{k}$. The vector $\mathbf{S}=\mathbf{E}\times\mathbf{H}$ is known as the *Poynting vector*, in units of watts/square meter. Integrating $\mathbf{S}$ over a closed surface gives the amount of energy per unit time (joules/second) flowing out of that surface.

Theoretical Basis of Metamaterials

Veselago's Model: In a seminal paper published in the late 1960s, the Russian physicist V. G. Veselago theoretically investigated the effects on EM fields of negative values of relative permittivity and permeability [3]. Because the refractive index is equal to the square root of their product, changing the signs of *both* ε_r and μ_r produces no change in the refractive index. Veselago suggested three possible interpretations of what simultaneously negative values of ε_r and μ_r might mean:

(1) The electromagnetic properties of any material are independent of whether or not ε_r and μ_r are both positive or both negative.

(2) Having simultaneously negative values for ε_r and μ_r is inconsistent with the laws of physics, and no such material can ever exist.

(3) Materials with negative permittivity and permeability may exist and possess unusual electromagnetic properties.

Veselago acknowledged that, at the time of his paper, there were no materials, naturally occurring or artificially made, with $\varepsilon_r<0$ and $\mu_r<0$. He nevertheless proceeded to analyze the properties of such media in the event that they might be discovered or man-made.

In the case of a monochromatic plane wave described by the propagation factor $e^{-i(kz-\omega t)}$, Maxwell's equations lead to the following relationships:

$$\mathbf{k}\times\mathbf{E}=\frac{\omega}{c}\mu_r\mu_o\mathbf{H} \tag{7}$$

$$\mathbf{k}\times\mathbf{H}=-\frac{\omega}{c}\varepsilon_r\varepsilon_o\mathbf{E} \tag{8}$$

If $\varepsilon_r > 0$ and $\mu_r > 0$, then the triad $\mathbf{E}$, $\mathbf{H}$, and $\mathbf{k}$ are related by the right hand rule. The medium is therefore *normal* or *right-handed* (RH). If, instead, $\varepsilon_r < 0$ and $\mu_r < 0$, the sign of the right hand side of equations (7) and (8) changes. As a result, the vector triplet is now left-handed (curling the fingers of the *left* hand). The direction of the wave vector $\mathbf{k}$ therefore is reversed when the material parameters are both negative. A metamaterial with negative permittivity and permeability is consequently called a *left- handed* (LH) material (LHM).

Power Flow in LHM: Compared to its orientation in RHM, the wave vector $\mathbf{k}$ in LHM points in the opposite direction relative to $\mathbf{E}$ and $\mathbf{H}$. An immediate consequence is that the Poynting vector and wave vector are in opposite directions. Because the Poynting vector $\mathbf{S}=\mathbf{E}\times\mathbf{H}$ always forms a right-handed triad with $\mathbf{E}$ and $\mathbf{H}$, regardless of the sign of the permittivity and permeability, allowing $\varepsilon_r < 0$ and $\mu_r < 0$ changes the direction of $\mathbf{k}$ but not of $\mathbf{S}$.

The wave propagates in one direction, but the power flow is in the opposite direction. Veselago points out that this situation corresponds to a negative group velocity ($V_{group}=\frac{\partial\omega}{\partial k}$, computed at the average value of the wavenumber). While perhaps somewhat surprising, there are examples of just this effect. Some naturally occurring anisotropic media and certain media in which spatial dispersion exists exhibit negative group velocity.

Doppler Effect and Cherenkov Radiation in LHM: Other unusual effects occur in LHM. The Doppler effect is reversed. The measured frequency of an approaching emitter will be lower than the stationary frequency, not higher as in RHM. Cherenkov radiation, electromagnetic radiation emitted by a charged particle moving at high speed in a LHM, will be directed backward relative to the direction of motion, which is the opposite of what happens in RHM.

Reflection and Refraction at a Plane Interface: EM plane waves incident upon a plane material surface are *refracted* as they enter the medium. For waves propagating between normal media, the wave bends towards the normal upon entering a denser medium (higher refractive index), and away upon entering a less dense medium. Figure 10 shows this effect in which the upper medium is less dense than the lower one. Φ is

the incidence angle, and Ψ the refraction angle. If the wave propagates from a RHM into a denser LHM, the refracted wave is bent towards the normal, but on the opposite side of the normal as shown in Figure 2. Interestingly, the direction of the reflected wave is the same (reflection angle equals the incidence angle) regardless of the nature of the lower medium.

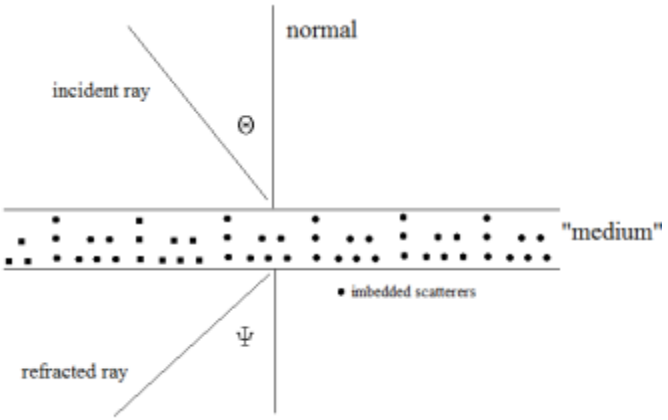


Figure 10. Refraction at MTM interface

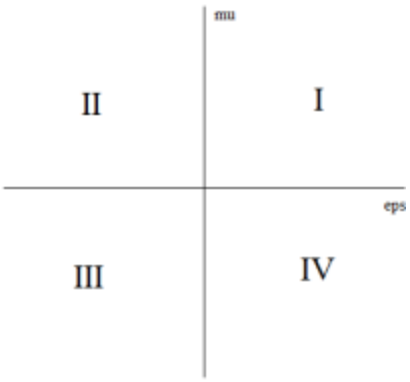


Figure 11. Media classification

Media Classification: In order to classify media based on permittivity and permeability, Veselago introduced the $\varepsilon - \mu$ diagram shown in Figure 11. The plane is divided into the four standard quadrants, labeled I through IV proceeding counter-clockwise from the upper right-most quadrant. Considering only isotropic materials, normal dielectrics with $\varepsilon_r > 0$ and $\mu_r > 0$ are contained in quadrant I. Quadrant II contains media with $\varepsilon_r < 0$ and $\mu_r > 0$, which is characteristic of gaseous and solid-state plasmas. The dielectric constant of a gaseous plasma is given by $\varepsilon = 1 - \sum \frac{\omega_o^2}{\omega^2}$, where ω_0 is the plasma frequency, and the summation is over all species in the plasma. At sufficiently low frequencies, the second term in this expression becomes greater than 1, which leads to a negative permittivity. Quadrants II and IV, however, are unoccupied, at least by isotropic materials, because there is no known naturally occurring material with negative permeability.

The $\varepsilon - \mu$ diagram can be used only for scalar values of permittivity and permeability. Certain anisotropic materials with tensor permittivity and permeability fall into quadrants III and IV for propagation in specific directions within the medium, because in those directions ε_r and μ_r are scalars. An example is propagation along the magnetic field direction in a gyrotropic medium, such as a plasma in a magnetic field. In

the propagation direction ε_r and μ_r are scalars, while generally in the medium they are second-rank tensors (3x3 matrices).

Requirement of Dispersion: A non-dispersive LHM cannot exist. This result follows from the fact that the energy density of an electromagnetic field is given by $W = \varepsilon_r \varepsilon_o \mathbf{E} \cdot \mathbf{E} + \mu_r \mu_o \mathbf{H} \cdot \mathbf{H}$. If both ε_r and μ_r are negative, then the total energy is negative, which is not possible.

In order to assure positive total energy, the medium must be *dispersive*. In a dispersive material, the expression for energy density is different, and includes the variation of permittivity and permeability with frequency:

$$W = \varepsilon_o \frac{\partial(\varepsilon_r \omega)}{\partial \omega} E^2 + \mu_o \frac{\partial(\mu_r \omega)}{\partial \omega} H^2 . \qquad (9)$$

W is always positive if $\frac{\partial(\varepsilon_r \omega)}{\partial \omega} > 0$ and $\frac{\partial(\mu_r \omega)}{\partial \omega} > 0$. In general, therefore, the permittivity and permeability must be frequency-dependent in LHM. Media in which both ε_r and μ_r are negative and constant with respect to frequency cannot exist. It does appear possible to have only one parameter depend on frequency, as long as its derivative is large enough to avoid a negative total energy; but Veselago did not consider this possibility.

The same conclusion about the necessity of dispersion can also be reached by considering the limiting case of a parallel plate capacitor at DC. Neglecting fringing, the capacitance is $C = \frac{\varepsilon_r \varepsilon_o A}{d}$, where A is the plate area, and d is the plate separation in consistent units. Physically, capacitance is the ratio of free charge on the plates, Q, to the potential difference (voltage) between them, V, taken as positive quantities. Thus, $C = \frac{Q}{V}$, must be positive at DC. If a metamaterial were used as a dielectric between the capacitor plates, its permittivity therefore must be positive in the limit of zero frequency (DC). This is necessary to avoid the physically impossible case of a negative capacitance. But, by definition, ε_r in a metamaterial must be negative at some frequency in order for the medium to be a metamaterial. Consequently, the metamaterial's relative permittivity must change sign with frequency, from a positive value at DC to a negative value at some non-zero frequency. This is dispersion. Note that this brief discussion does not address how the capacitor behaves as a circuit element at frequencies where $\varepsilon_r < 0$. Doing so should be quite interesting.

Realization of Metamaterials

Because metamaterials do not occur naturally, the only way to obtain a metamaterial is to manufacture it. The very nomenclature, *metamaterial*, conjures up the image of a "substance", generally a solid material with some specific chemical composition. Existing metamaterials, however, could not be further from that image.

Smith [8] recently described a LHM that was modeled theoretically and measured experimentally. Building upon the split ring resonator (SRR) medium created by Pendry [9], short, small-diameter wires were added to a an array of SRRs to create the left-handed metamaterial. The medium is actually an array of discrete electromagnetic scatterers arranged so that the *effective* ε_r and μ_r are negative at some frequencies. It is not at all a "substance" in the usual sense.

Figure 10 represents this concept schematically. Imagine two parallel planes in space, between which an array of scattering elements is deployed. The scatterers, represented by dots in the figure, might be short wire segments, other geometrical shapes, SRRs, other resonant structures, or any number of EM elements fabricated from conductors and/or dielectrics.

In the realm of every possible configuration, it seems intuitively plausible that some arrangements will result in anomalous propagation such as seen in metamaterials. In Figure 10, for example, it is not difficult to imagine that there is some scatterer arrangement that will "refract" a wavefront incident at the angle Φ in the direction opposite to "normal" refraction as shown. This type of propagation is a characteristic of a left-handed material. If the volume between the planes is thought of as a "medium", it would be properly described as one with negative effective permittivity and permeability, that is, a metamaterial.

Metamaterials that are "substances" may eventually be developed or discovered. But current research focuses on creating LHMs using imbedded scatterers. This is not a limitation, however, and indeed may actually provide more flexibility than attempting to create purely chemical substances. Designing the scattering matrix may permit engineering metamaterials whose properties could not exist in chemical substances. Creating solid materials consisting of a matrix containing small conductive scatterers, possibly even carbon nanotube scatterers, might permit the designer to actually tailor the LHM properties to the EM application.

The idea of using a matrix of scatterers to create a "medium" with a negative effective permittivity is not new. In the 1950s, Bracewell [10] and Rotman [11] experimentally investigated ionospheric propagation using a three-dimensional array of

intersecting thin, straight wires. Dispersion in this "medium" is similar to that in a neutral plasma, such as the ionosphere.

Just as the real ionosphere does, the artificial one has $\varepsilon_r < 0$ for frequencies below the "plasma frequency". However, unlike the ionosphere's plasma frequency, which depends on the charged particle density, and particle charge and mass, the artificial medium's "plasma frequency" depends on the arrangement and dimensions of the wires. Because LHM must exhibit $\varepsilon_r < 0$ at some frequencies, and because the dispersion relation has the same form as a gaseous plasma's, metamaterials made from scatterers are often described in terms of a *plasma frequency*. It is important to remember that the term used in the manner is a purely formal construct, not to be confused with the parameter of the same name describing a gaseous plasma.

The effective permittivity of Smith's wire medium is given by

$$\varepsilon_{eff}(\omega) = 1 - \frac{\omega_p^2}{\omega^2} \qquad (10)$$

where ω_p is that medium's plasma frequency. ω_p depends in a complicated way on the wire scatterers' electrical properties and geometrical arrangement. At frequencies below ω_p, ε_{eff} is negative, which is one of the requirements for LHM. The other requirement is that μ_r simultaneously be negative. The SRR medium has an effective permeability given by

$$\mu_{eff}(\omega) = 1 - \frac{F\omega_o^2}{\omega^2 - \omega_o^2 - i\omega\Gamma} \qquad (11)$$

where F is the fractional area of the unit cell occupied by the SRR's interior, and Γ is the SRR dissipation factor. ω_o is the SRR medium's plasma frequency. Equations (10) and (11) assume that the array of wires and the array of SRRs do not interact appreciably, which was true in Smith's medium.

When the scatterer arrays are uncoupled, the wire array produces the medium's ε_{eff}, and the SRR array produces its μ_{eff}. Under these conditions, the effective refractive index of the composite medium is $n_{eff}(\omega) = \sqrt{\varepsilon_{eff}(\omega)\mu_{eff}(\omega)}$. If there is an overlapping band of frequencies in which the effective permittivity and permeability are both negative, then the effective index of refraction is real, and propagating modes exist. The medium is LH, however, because of the negative material parameters, so that the unusual propagation effects associated with LHM occur.

The SRR, wire segment, wire-SRR media were modeled using a commercial EM software package to generate dispersion curves (the problem is far too complex for an analytical solution). The SRR array alone showed a pronounced band gap from 4.2-4.6 Ghz. No modes would propagate in this band. The wire array alone exhibited an even wider band gap, from just above DC to 13 GHz, the array's plasma frequency. No modes would propagate in this range. The reason propagation cannot occur is that μ_{eff} and ε_{eff}, respectively, are separately negative in these bands, which results in a purely imaginary refractive index.

A metamaterial is created by combining the wire and SRR arrays. The wire-SRR array has overlapping bands from 4.2-4.6 GHz where the simultaneously negative material parameters combine to create a metamaterial. Computer modeling of the dispersion characteristics of the combined arrays clearly showed propagation in the otherwise forbidden band. Experimental measurements further confirmed the LH nature of the medium by observing propagation between 4.2 and 4.6 GHz which was not possible in either array alone.

Analysis of Metamaterials

In calculating a metamaterial's performance, it is unlikely that analytical results are attainable for all but the simplest scatterer geometries. Analytically solving Maxwell's equations is almost certainly impractical. Computer modeling is likely to be the best approach, and possibly the only truly useful one. Metamaterials made of imbedded arrays of scattering elements can, in fact, be analyzed with existing modeling software, at least to some degree. Commercially available packages, and government-developed programs such as NEC-4, provide a starting point for modeling metamaterials. But many questions arise.

Most EM software approximates the modeled structure using some form of discretization. In NEC-4, for example, continuous wires are approximated by short segments. Continuous surfaces are approximated by patches of surface area. Other programs use the same or similar approaches.

One characteristic of the discretized models is that very substantial computer resources (memory and speed) are often required to obtain good results. In NEC-4, for example, the model size (a measure of required memory) is proportional to the square of the number of segments, and the computation time (a measure of required processor speed) scales similarly. Other types of models exhibit similar characteristics. Disproportionately greater resources are required with increasing complexity, modeling accuracy, and resolution in the frequency domain. It may be expected that these issues will be exacerbated for metamaterials, because LHMs fabricated from large numbers of

complex individual scatterers are indeed very complex structures. This level of complexity is likely to push existing modeling capabilities to the limit.

Metamaterials should be studied experimentally, not only to characterize available materials, but also to develop the experimental techniques and fixtures needed for such studies. Because of the anomalous propagation, it is likely that new measurement techniques will have to be developed to fully explore metamaterial properties. For example, direct measurement of the negative values of permittivity and permeability, key parameters in characterizing a metamaterial, is required. But it is not clear that suitable instrumentation or techniques for doing so exist.

Future of MTM

MTM hold great promise as components in antenna design, particularly at microwave frequencies. Whether or not they can be used effectively a low VHF frequencies is an open question because there simply is not enough data to reach a conclusion. At this time not enough is known about MTM's electromagnetic behavior. There are no "design handbooks" and essentially no low frequency data. Nevertheless, the performance of MTM-based antennas at microwave frequencies strongly suggests hat MTM antennas will be useful at VHF as well.

2.7.7 References

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2.8 Electronic Band-Gap and High Impedance Surfaces

Introduction

Electromagnetic band-gap (EBG) structures recently have attracted a great deal of interest among researchers because EBG structures can significantly reduce surface waves, thereby improving patch antenna performance [1]. Reflection phase characteristics also can have a pronounced negative effect on antenna performance [2,3]. It is well known that when a dipole antenna is operated near a metallic ground plane as shown in Figure 1, the radiated waves are reflected from the ground plane, so that interference may occur. Antenna bandwidth and gain are degraded, which is readily demonstrated by applying image theory.

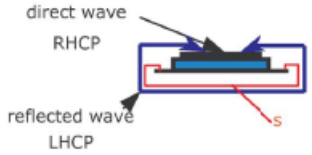

Figure 1. Antenna Near Ground Plane

To overcome these difficulties, dipoles are operated at a height of 0.2% or higher above the ground plane. In [1] the concept and feasibility of developing wideband, ultra-thin (substrate thickness < O.1) directional dipoles and dipole arrays using EBG structures are addressed. The reflection phase angle of an EBG surface varies from -180 to 180 degrees with frequency. The goal is to make the antenna operating frequency band coincide with the EBG stop band, and then to flatten the reflection phase curve within the stop-band region, which then will ensure wide-band operation of the antenna. This technology could be useful in an integral television antenna as well as set-top antennas that lay flat on the television.

The parameters generally adopted to characterize the performance of EBG structures include the surface wave attenuation and reflectivity. A double benefit can be derived from the use of artificial magnetic conductor type of EBGs because not only are surface waves suppressed, but also in-phase image currents are induced. The so-called reflection phase—frequency diagram gives information on how the structure reacts to a wave impinging on it. Normally, the configuration behaves similar to a metal plate (perfect electric conductor, “PEC”). However, another characteristic feature of EBGs is the existence of a certain frequency range over which an incident electromagnetic wave does not experience *any* phase reversal upon reflection (a rather remarkable phenomenon). In this range the structure behaves as a perfect magnetic conductor (“PMC”). The frequency range where the phase reversal is within from the PMC point (phase reversal) is conventionally used to define the bandwidth and is called the “bandgap.” The concept here is to introduce a ground plane for shielding purposes (reducing the contributions of LHCP signals), and cover this plate with EBG cells (cancelling out the surface currents)

Types of EGB-Cells

Various EBG structures from literature have been investigated. A small sample of different geometries is displayed in Figure 2. These EBG cells can be with or without vias to a ground plane on the opposite side of the substrate. A Mushroom EBG substrate is shown in figure 3. The EBG cells in this figure are shown with vias (each cell is shorted to the ground plane).

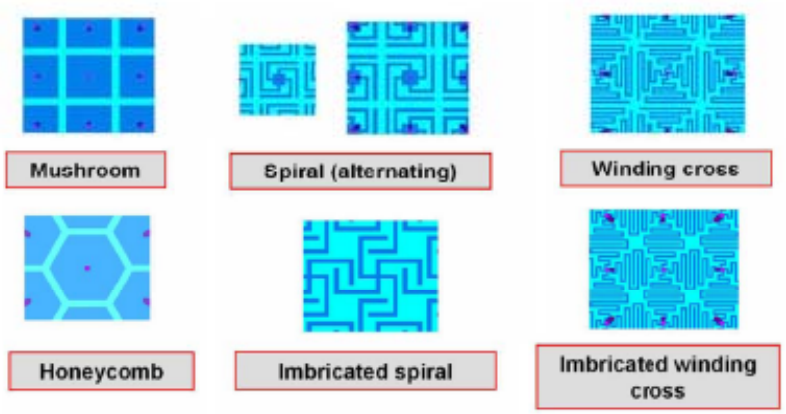

Figure 2. Different Types of EBG Structures.

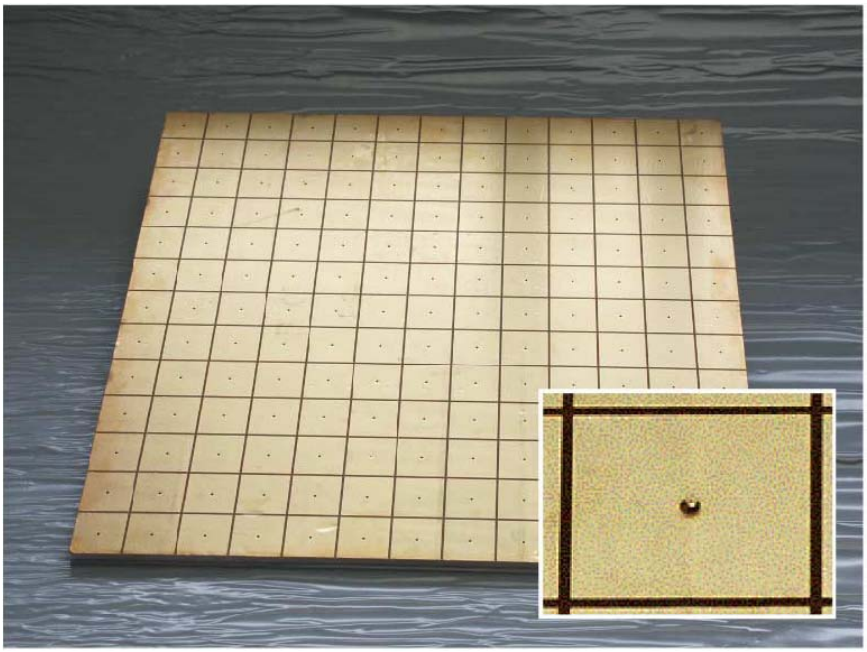

Figure 3 Mushroom EBG substrate (inset: Zoom on a single cell and a golden nail)

Performance of a Dipole Antenna on a Constant Reflection Phase EBG Surface

For a dipole antenna operating against a ground plane, the driving-point impedance is the algebraic sum of the self and mutual impedances when image theory is applied. For a PEC ground plane the driving point impedance is the self minus the mutual impedance since the reflection phase is 180 degrees. The driving point impedance of a dipole on an EBG structure with any other constant reflection phase characteristics can also be

calculated. For example, for a thin wire dipole the self and mutual impedances can be computed using the induced EMF theory. Of course, it may not be possible to achieve a constant reflection phase over a band of frequencies using an EBG structure, but that is the goal. As an example, the dipole under consideration is infinitely thin and resonant at 2.9 GHz in free-space with a 2:1 VSWR bandwidth of *5%.* The driving point resistance is negative (at low frequencies) for phase angles of -100 degrees to -160 degrees because the mutual resistance is larger than the self resistance and adds destructively with it. But for phase angles of 40 degrees to 110 degrees the mutual resistance adds constructively with the self-resistance. Thus, the driving point reactance for different phase angles is capacitive at low frequencies and inductive at high frequencies.

An EBG surface can be characterized by its reflection phase angle given by

$$\phi(f) = \mathrm{Im}\left[\ln\left(\frac{af^2 + bf - c}{af^2 + bf + c}\right)\right]$$

where a, b and c are the EBG design parameters with $a=4\pi^2\eta LC$, $b=2\pi L$ and $c=\eta$ [4]**.** Here ***L C*** and η represent the inductance, capacitance and intrinsic impedance of free-space, respectively. The slope of the reflection phase curve can be changed by varying the values of a and b*.* The bandwidth of this dipole antenna has been enhanced by flattening out the reflection phase curve at the useful phase angle regions, which corresponds to the EBG surface of having a stop band center frequency at 2.82 GHz.

EBG Dipole

Work on an EBG Dipole was completed in the past by MegaWave Corp. Figure 4 shows an EBG structure modeled in XFDTD. It consists of a double layer, mushroom-like structure embedded in FR-4 substrate with the bottom layer consisting of an array of 6 by 12 square patches and the top layer of an array of 5 by 10 square patches. The overall dimensions are 2”x4” with a thickness of 0.18”. The computed reflected phase of the EBG is plotted in Figure 5. The reflected phase was generated by comparing the phase of a plane wave scattered from the EBG surface with that from a PEC surface. The bandwidth of the EBG material was estimated to be from 1.66 to 2.35GHz as determined by the reflection phase being in the range of 90°±45°[2].

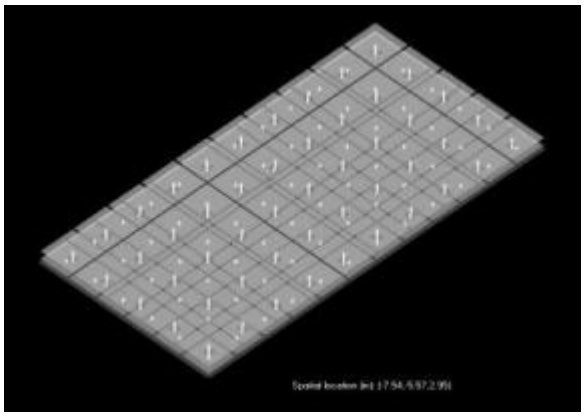

Figure 4: XFDTD Model of Dual Layer EBG Material

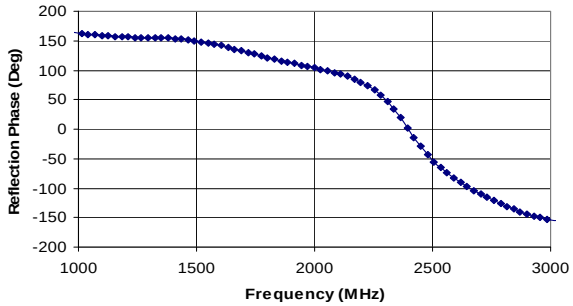

Figure 5: Computed Reflection Phase of EBG Material

A prototype of the MegaWave-designed EBG structure was fabricated as shown in Figure 6, and Figure 7 shows a wideband dipole placed 0.25" away from the EBG structure. The measured VSWR was less than 2.6:1 across a very wide operating frequency band from 1.35 to 2.7GHz as shown in Figure 8.

Figure 6: Lab Prototype of EBG Structure (Top and Bottom Layers Shown)

Figure 7: Lab Prototype of Wideband Dipole on EBG Material

The free space radiation patterns of both the wideband dipole on the EBG structure and the Phase I directional dipole antenna were measured and compared in MegaWave's laboratory. Figures 9 and 10 show the measured free-space patterns. This EBG technology could be implemented in the television UHF and VHF bands with an expectation of greatly improved set-top antenna performance.

Figure 8: VSWR of Wideband Dipole on EBG Material

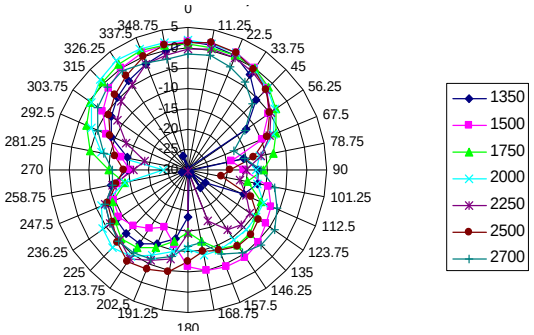

Figure 9: Measured Radiation Patterns of Antenna

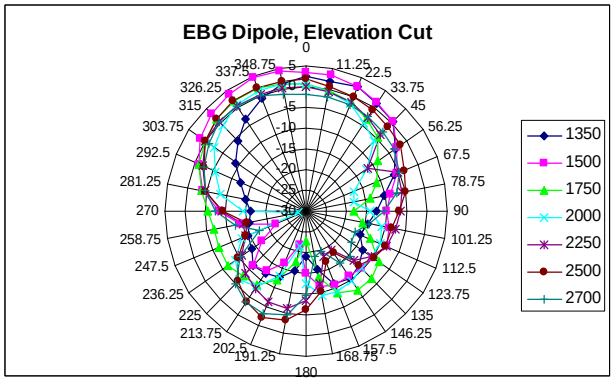

Figure 10: Measured Radiation Patterns of Wideband Dipole on EBG

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2.9 Fractal and Self-Similar Antennas

There is no question that, among all the antennas existing today, one of the most interesting and visually appealing is the Fractal antenna. However it not clear exactly what benefits in bandwidth and gain they provide as compared to non-fractical elements of the same shape and area, but without the fine detailed geometries. The reader is directed to the section on pathological antennas (1.5) and its subsection 1.5.8 on fractals in Dr. R.C. Hansen's recent textbook, "Electrically Small, Superdirective, and Superconducting Antennas," Wiley Interscience, 2006.

First publications about electrodynamic characteristics of fractal structures were published in the 1980's, but first practical approach appeared 10 years later. Dr. Nathan Cohen, a professor at Boston University, designed, engineered and patented many fractal antenna designs, some of which are shown in Figure 1.

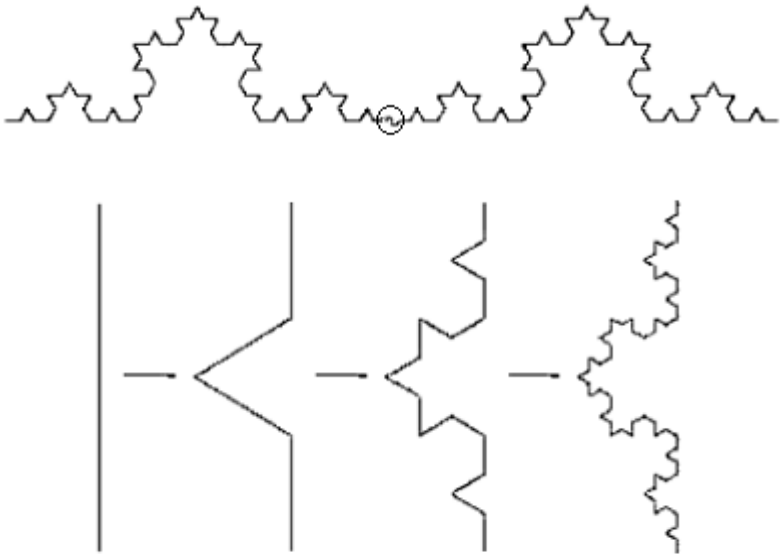

Figure 1. Fractal Designs

This fractal figure is built as follows: first line (length is z) is cut in 3 equal pieces z/3. In the middle the triangle is formed with same side lengths z/3 and same angles. This creates a single element template. This process is repeated with other segments in which sizes diminish 3 times (z/9), then again 3 times (z/27), and so on. A fractal antenna is created using fractal geometry, a self-similar pattern built from the repetition of a simple shape.

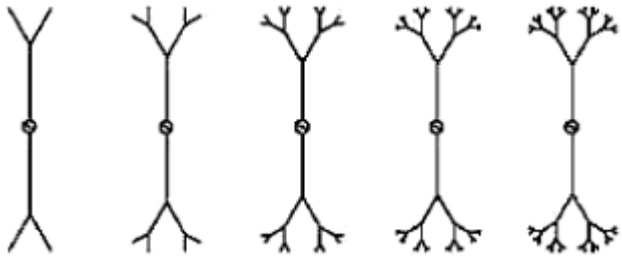

Figure 2. Fractal Dipole

The Fractal dipole as shown in Figure 2, is a recursive tree. Each of the new iterations multiplies the branches by two, in the process lowering the resonance frequency. By using many iterations, it is possible to use an antenna at lower frequencies without increasing the overall antenna size. Dipole antennas usually have narrow bandwidth, about 2.4% of the center frequency. If the 5th fractal iteration is used, then this parameter increases to 3.1%. If a 3D tree is used (when there are 4 branches used instead of 2) then this parameter increases 12.7%. Besides Dipole antennas there are resonant loop antennas used. Some typical loop fractal antennas built using Koch figures as shown in Figure 3 [7,8].

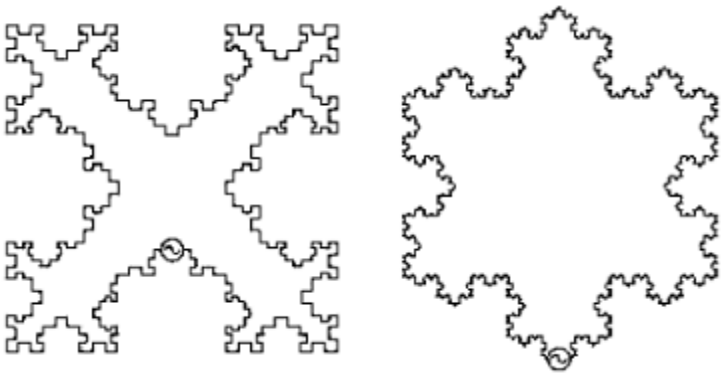

Figure 3. Fractal Loop Examples

RF MEMS could also be used in conjunction with fractal antenna structure as the basis of a new reconfigurable array antenna approach. The RF MEMS switches permit the connectivity of sections of the antenna's conductive parts, and therefore enhance the coupling between the fractal elements allowing multiple frequency operation with one antenna. Using this scheme a set-top antenna could be designed to change shape and/or size based upon which frequency (channel) the television was currently tuned to. The use of fractal shapes provides the multiband characteristic through the property of self similarity at equal or different physical scales. The electromagnetic performance of the RF MEMS switches is ideal for this application, and their placement is accomplished by small physical connections of the antenna's adjacent conducting pans [1].

Most fractal antennas, including the Sierpinski antenna, have been studied extensively over the last few years [3,4]. A modified Sierpinski antenna is shown in Figure 4. The centerline of the CPW [1] feed provides the RF input and dc voltage for MEMS switch actuation. Switches S1 and S1' actuate at a low voltage, switches S2 and S2' at a medium voltage, and switches S3 and S3' at a high voltage. Figure 5 shows the different states of the antenna. An alternate implementation could be for a bowtie antenna as shown in Figure 7.

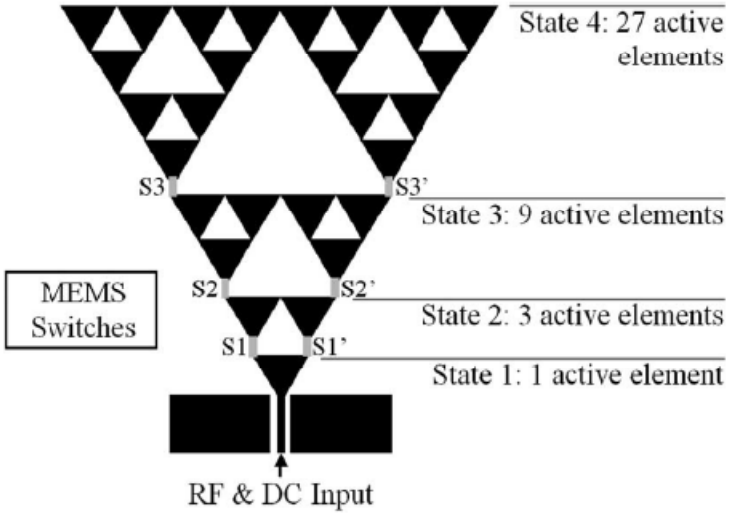

Figure 4. Illustration of a MEMS reconfigurable Sierpinski antenna.

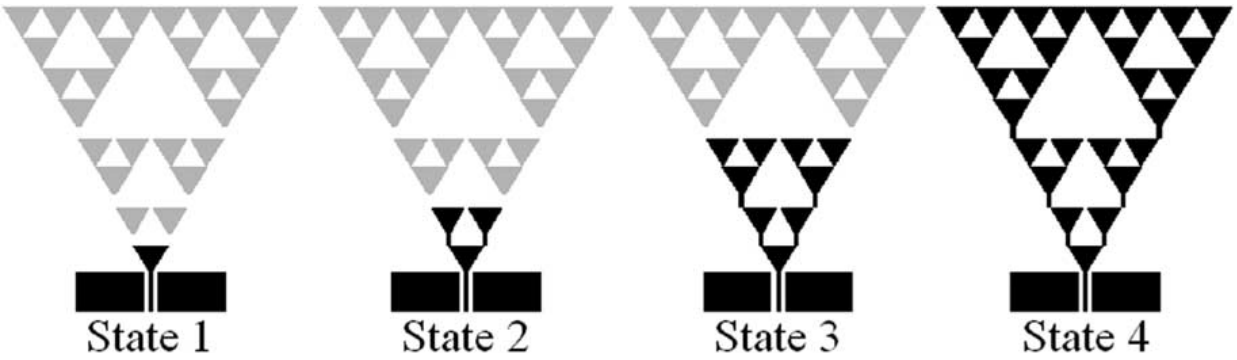

Figure 5. States of Reconfigurable Sierpinski Antenna

In most of the literature, integration has been accomplished on rigid and nonflexible semiconducting or organic polymer substrates such as silicon and FR-4. The idea of integrating RF MEMS switches into a multiband self-similar antenna was first implemented in [2], where the entire system, including the RF MEMS, the planar self-similar antenna, and the CPW–coplanar strip transition, was fabricated on silicon. The integration is achieved on a very thin and flexible liquid crystal polymer (LCP) substrate. Since the switches share a common dc feed, this technique provides reconfigurability

without the need for additional bias lines. This is advantageous since dc bias lines take up space, add loss, and reduce the bandwidth of a device. This technology is particularly useful for antennas where bias lines can have a pronounced effect on radiation patterns, and may be especially useful in the set-top application.

The implementation of a sequentially activated antenna is shown in Figure 5. All of the MEMS switches used are single supported (cantilever type) and ohmic. Regardless of the applied voltage, the triangular element that is closest to the RF/dc input is always active. When no dc voltage is applied, the antenna radiates at its highest frequency. When a low dc voltage is applied to the signal line, the first set of MEMS switches (S1 and S1') actuate, and this activates the second level of triangular elements. The antenna now radiates at a lower frequency. Since all of the switches are ohmic, the low voltage is also present at the membrane of the next set of switches (S2 and S2'). However, these switches are designed to actuate at a higher voltage so they are unaffected by the voltage present. When a higher dc voltage is applied, the first set of MEMS switches (S1 and S1') remains in the "ON" position while the second set of switches (S2 and S2') actuates. This activates the next iteration, consisting of six additional radiating elements. Again, this higher voltage is present at the next set of switch membranes (S3 and S3'), but the electrostatic force created is not sufficient for actuation. Finally, when the voltage is increased to its highest value, the first two sets of switches (S1 and S1' and S2 and S2') remain in the "ON" position while the remaining set of the switches (S3 and S3') actuates .

Thus, the voltage "cascades" from one state to the next like a sequence of overflowing buckets. This technique could not be used if the switches were capacitive since they do not pass dc voltage. The four different states are illustrated in Figure 8, where all of the activated regions for different voltages are dark in color. This biasing technique allows for direct actuation of the electrostatic MEMS switches through the RF feed structure. Since only the RF feed is dc grounded, the switches actuate with the use of a floating ground. That is, the signal pin of the CPW feed is connected to the dc cathode, and the ground pins are corrected to the anode of a bias tee. The electrostatic charges that are created during switch actuation can dissipate through the substrate and be removed by the dc-grounded RF feed when in the "OFF" state. This method has been successfully documented in [5]. The reduction or elimination of bias lines is highly advantageous because they can significantly distort the radiation patterns and they can introduce additional unwanted resonances. To change the actuation voltage of a MEMS switch, there are four parameters that can be changed.

1) *Membrane material:* Switch membranes are almost always made of metal. This is due to their pliable nature. Stiffer metals (that is, those with a high Young's Modulus*, E*) will have a higher actuation voltage than those with a lower *E*.
2) *Bridge thickness:* The thicker the bridge, the stiffer the membrane, resulting in a higher actuation voltage.

3) *Bridge height:* The higher the bridge, the larger the gap between the metal layers, resulting in decreased the electrostatic force and increased actuation voltage.
4) *Membrane geometry:* Springs can be designed into the shape of the membrane to lower the actuation voltage.

Of these parameters, only the fourth does not require any fabrication changes. Making changes to a fabrication process can be a costly endeavor and may create additional variables of concern.

LCP was chosen as the substrate for its many advantages. LCP is a thin (100 μm), flexible, low-loss (tan $\delta \approx 0.004$), low-moisture absorbing material with low permittivity ($\varepsilon r \approx 3$) [9]. Since the material is a polymer, there are additional packaging and cost advantages. All of these characteristics make it an ideal substrate for antennas, particularly at high frequencies. With respect to the geometry, the antenna elements have a 60◦ flare angle and maintain the resonant structure's self similarity with a log-periodicity of $\delta = 2$. The antenna is fed through a 6-mm-long CPW transmission line with a 50-μm gap, a 1.3-mm signal conductor width, and a 1.5-μm-thick aluminum layer. The overall size of the antenna, including the feed, is 20 × 25 mm. The CPW feed was chosen to facilitate the measurement setup. This reconfigurable antenna operates at four different principle frequencies. For each frequency, the antenna maintains its multiband performance. It was verified that the antenna has a different first resonant frequency for each of the four states. Since the antenna is self-similar with a log-periodicity of 2, each time the antenna transitions to the next state, the frequency should be halved. That is, the resonant frequency of state 2 should be half that of state 1. This is shown in Figure 6.

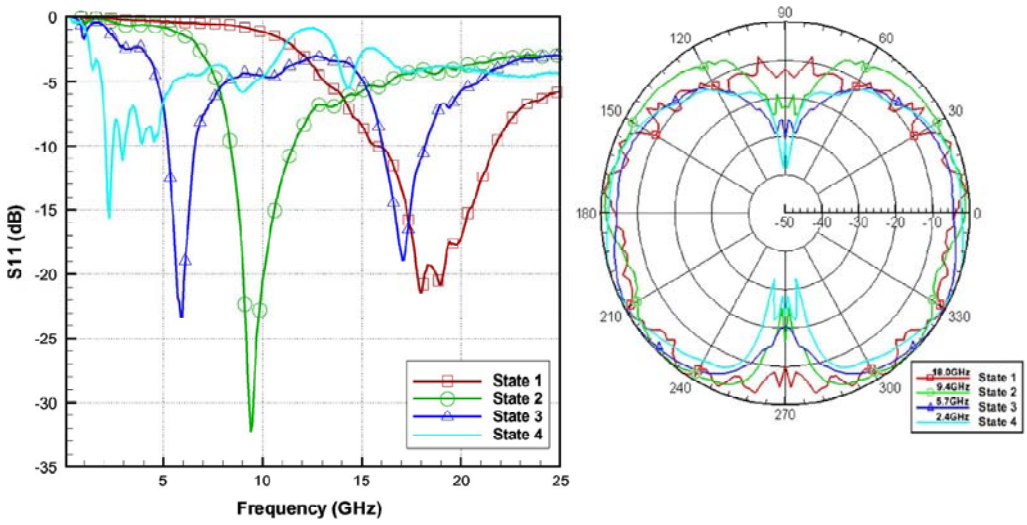

Figure 6. Measured Results of Built antenna for S11 and pattern

Fractal Antenna with RFID-MEM's

The general problem with using MEMS in a fractal antenna is how to power them. This was accomplished by putting DC on the RF of the feed of the antenna with a floating ground in the last section. If the goal is a dipole antenna as shown in Figure 7, then the floating ground would become a problem. One possible solution for this difficulty is known as "RFID-MEM's." The RFID could power the MEM to switch a section on and off. It would be controlled with a reader placed close to the antenna. If the television sends a digital signal back on the coaxial line that has information on what channel is currently tuned, then this information could be used by the reader to activate the correct switches to optimize the antenna for the best reception on the tuned channel. The RFID would need to operate either below the television spectrum, such as 13.56MHz, or above it, such as an UWB RFID. This system should have minimal effects on the antenna pattern of the bowtie antenna and could cover the entire VHF and UHF bands.

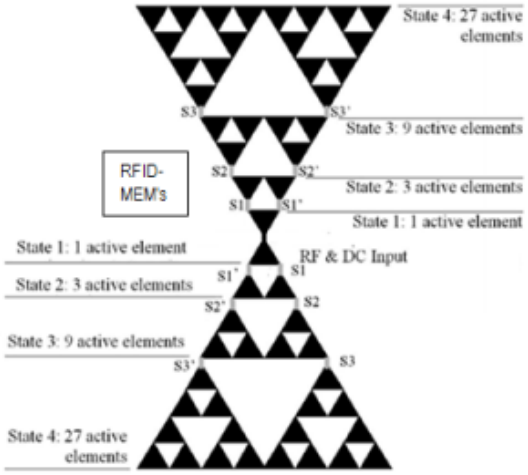

Figure 7. Bowtie Antenna

Fractennas

There have been some commercially available fractal/self similar antennas made by Fractal Antenna systems Inc. [10]. Fractal Antenna Systems has fabricated a wideband dipole as shown in Figure 8. The antenna has a self similar geometry present at multiple resolutions. Along two of the axes there is a substantial mirror image. Along the center there is a thin dipole that runs all the way through the antenna. The long thin dipole geometry creates a loading effect rather than predominately providing the function of radiating electromagnetic energy, which instead is provided by the more distinct V shaped dipole as shown in Figure 9. This antenna is claimed to radiate energy between 400 MHz and 6000MHz and are fabricated using deterministic fractal geometry.

Figure 8 Wideband dipole by Fractenna

Figure 9 Wideband Dipole (Bowtie) by Fractenna

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[10] Fractal Antenna Systems, Inc. 130 Third Avenue Waltham MA 02451, www.fractenna.com

2.10 Retrodirective Arrays

2.10.1 Summary

Retrodirective antennas are self-adapting systems that transmit a signal or direct their beam in the direction of an incoming signal without any prior knowledge of the signal's angle of arrival or intensive signal processing to determine the direction. Two types of retrodirective antennas have been developed, the Van Atta (VA) array and the Phase Conjugating (PC) array. The VA array is a fundamentally a passive, planar structure (although active VA arrays exist as well). The PC array is an active antenna that can be fabricated on a non-planar surface. Thus, the VA array generally is simpler and, consequently, easier to design and fabricate. Retrodirective systems are attractive candidates for the set-top television antenna.

While existing retrodirective systems typically operate at frequencies well above the television band, the technology is advancing rapidly and much of it at this time or in the short-term future likely will be applicable to set-top antennas. This section describes several state-of-the-art implementations, mostly for VA systems, but some for PC systems. It also describes an important development in conformal PC array design using genetic algorithm optimization that allows a retrodirective array to be designed for a specific surface shape. Finally, some new work on near field scatterer effects is discussed that shows how a retrodirective antenna can mitigate near field phase distortions in environments such as the interior of a building (a typical set-top environment).

(a) Retrodirective Concept

A retrodirective antenna is conceptually similar to an optical corner reflector, three plane reflecting surfaces intersecting at right angles. A light beam incident onto the corner reflector, also known as a retroreflector or cataphote, is automatically reflected back along the direction from which it came. The basic retrodirective array concept is shown schematically in Figure 1. There are two basic structures: (1) the Van Atta array, and (2) the Phase Conjugating array. The VA array is simpler than the PC array because it does not require active devices, whereas the PC array does. VA and PC arrays operate differently, but both systems achieve the same goal of directing the antenna's beam in the direction of an incoming signal. Without any knowledge of the signal's direction of arrival, a retrodirective antenna is able to transmit a signal back toward the transmitter along the direction of that incoming signal without using phase shifters.

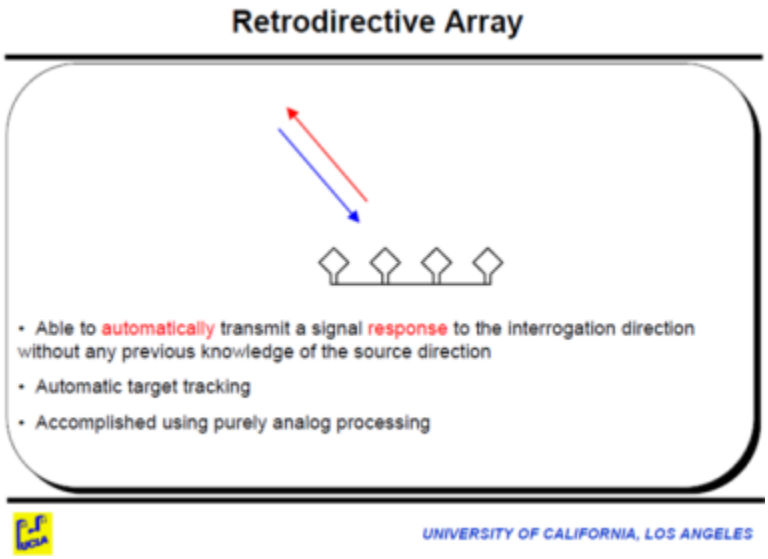

Figure 1. Basic retrodirective concept (reproduced from [1].

(b) Van Atta Array

The VA concept [2] is shown in Figure 2. Pairs of antenna elements, which can be any type of radiator (in this case horns), are equally spaced about the array's center point and interconnected with waveguides (or transmission lines) that have equal length or are multiples of a wavelength. Each antenna element acts as both a receiving antenna and a transmitting antenna. The incident field received by each element is reradiated by its paired counterpart after propagating along the interconnecting transmission line or waveguide. The effect of the resulting propagation delay is to reverse the phase of the reradiated field distribution relative to the received field distribution, causing the reradiated field to be directed back along the line of sight of the incident field. While Van Atta's original antenna was strictly a passive retroreflector, active systems have been developed that utilize the basic Van Atta structure [3,4,5]. Figure 3 provides a schematic illustration of the Van Atta concept for passive and active implementations.

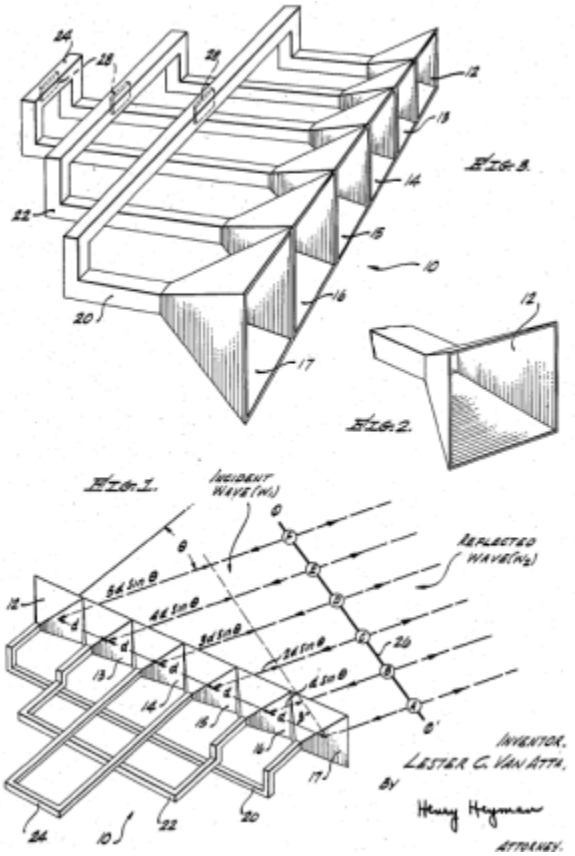

Figure 2. Van Atta retrodirective array (reproduced from [2]).

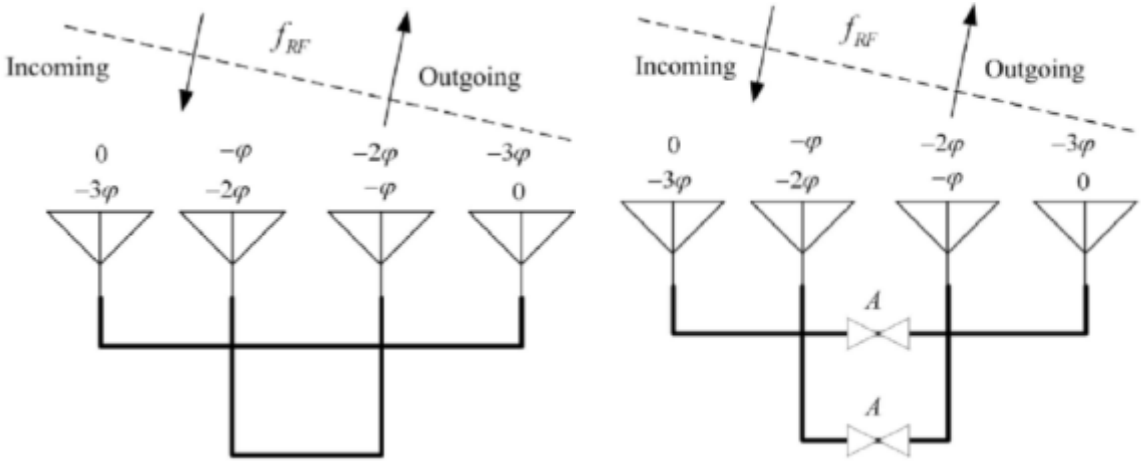

Figure 3. Passive (a) and active (b) Van Atta array phase front (reproduced from [3]).

(c) Phase-Conjugating Array

Unlike the Van Atta array, the PC retrodirective array requires active elements because its operation relies on actively changing the relative phase of its elements. The direction of an emitted signal, or, by reciprocity, the maximum directivity for received signals, is controlled by the relative phase and amplitude of the individual radiating sources or receiving elements in a multi-element array as shown in Figure 4. By operating a local oscillator (LO) at twice the received signal's radio frequency (RF), mixing the LO and RF signals results in a lower sideband product whose frequency is the same as the incident signal's but whose phase is conjugated (reflected about the real axis as shown in Figure 5).

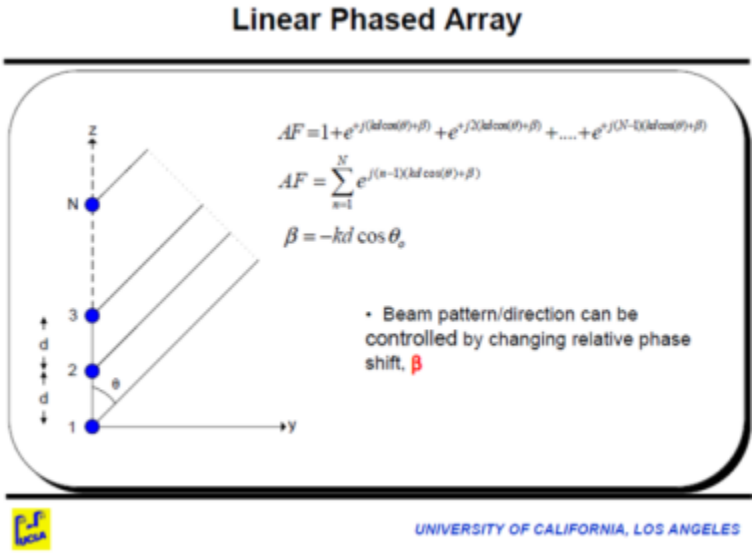

Figure 4. Beam direction control (reproduced from [1]).

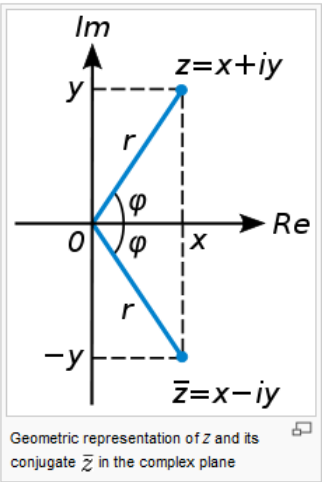

Figure 5. Phase conjugation.

Conjugating the phase for each antenna element steers the antenna beam back along the direction of the incoming signal as shown in Figure 6. If the antenna is used for transmitting, then the radiated signal is transmitted along back along the direction of the received signal. If the antenna is used for receiving, then the radiation pattern is steered with is maximum directivity in the direction of the received signal.

A conjugating retrodirective antenna, whether VA or PC, could be very useful as a set-top television antenna because it provides optimal gain in the direction of the received signal regardless of its incidence direction. A retro receive antenna would automatically adapt itself to the signal environment allowing a fixed antenna to provide maximum response to signals from any direction. The following sections describe some state-of-the-art implementations that may be applicable in the set-top television environment.

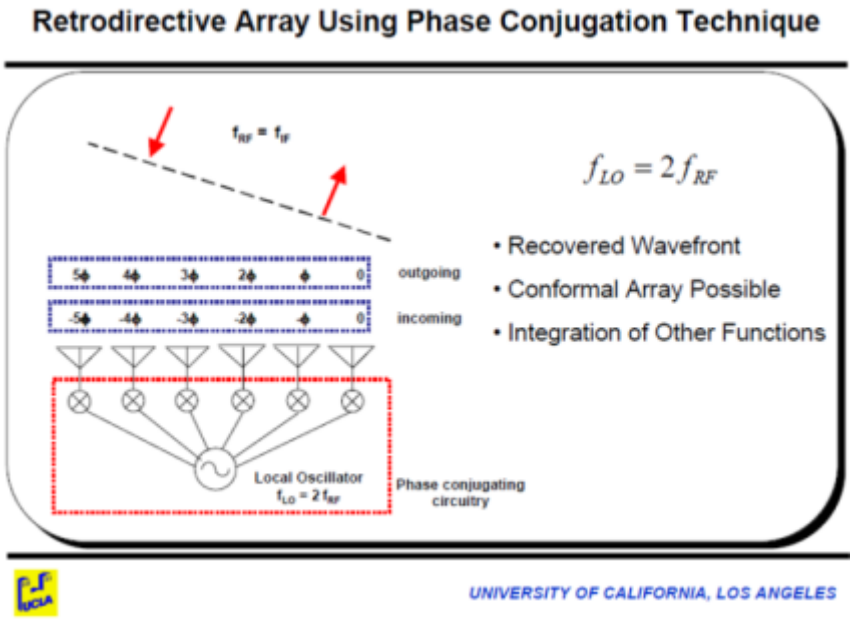

Figure 6. Retrodirective phase conjugating architecture (reproduced from [1]).

(d) Retrodirective Array Implementations

Broadband Van Atta Array. Figure 7 shows the schematic diagram for a frequency-offset broadband Van Atta array [6]. While the VA architecture is inherently more broadbanded than PC arrays because bandwidth-limiting LO and phase conjugation mixers are not required, there nevertheless are active-element implementations that improve performance while maintaining substantial bandwidth as shown in the figure. In this particular application, the antenna is used for two-way communication, which is why optional amplifiers that increase effective radiated power are included and dual-polarized antennas are used (orthogonal polarizations for transmission/reception to increase isolation).

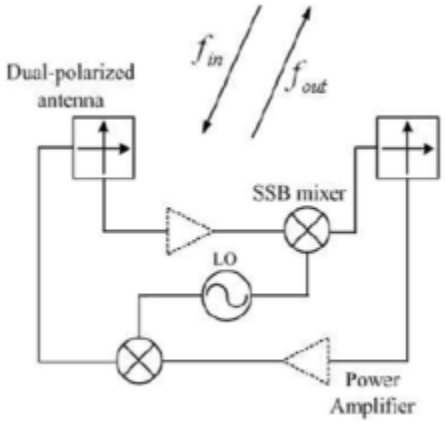

Figure 7. Diagram of broadband two-way Van Atta array (reproduced from [6]).

Figure 8 shows the structure of the prototype broadband VA array antenna. A patch radiator is fed by two excitation slots at right angles to provide orthogonal polarization. Microstrip feed lines excite the apertures and are deposited as shown the opposite side of the aperture substrate. The reflecting plane is included to improve front-to-back ratio. Two such antennas comprise the prototype array which operates from 2.8 to 3.4 GHz. The antenna as fabricated is shown in Figure 9. The patch antennas are evident on the surface of the front circuit board substrate, while the separate SSB mixer circuit boards are seen in the background. This antenna achieved beam steering over approximately ±30º from the broadside direction (perpendicular to the plane of the antenna patches). This active VA array shows that substantial retrodirectivity can be achieved even in a simple two-element array. It perhaps points the way towards retrodirective set-top antennas based on the same or similar architecture that utilize different antenna elements, such as dipoles or bow-ties, that are better suited to the lower television frequencies.

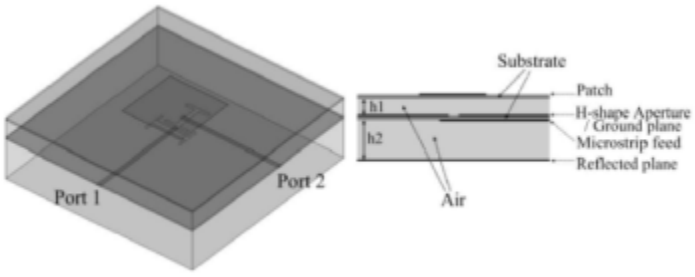

Figure 8. Broadband two-way Van Atta array antenna element (reproduced from [6]).

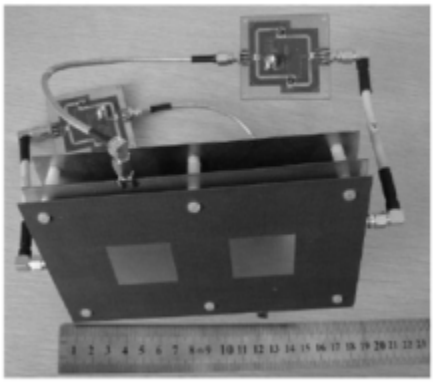

Figure 9. Broadband two-way Van Atta array as fabricated (reproduced from [6]).

Passive Array with Hybrids. An interesting example of a passive VA array implementation involves using 90º hybrid couplers as the phasing elements [7]. This design relies on a clever use of hybrid couplers whose incident and reflected signals' phases are reversed when the coupler is terminated with loads having the same reflection coefficient. Figure 10 shows how a phase difference $\Delta\phi$ between signals into ports 1 and 2 (input/isolated ports, respectively) is transformed into a phase difference of $-\Delta\phi$ between the reflected signals at those ports as long as ports 3 and 4 (direct/coupled, respectively) are terminated with the same reflection coefficient. In practice, these ports simply float unterminated because Γ then is the same for each (~1).

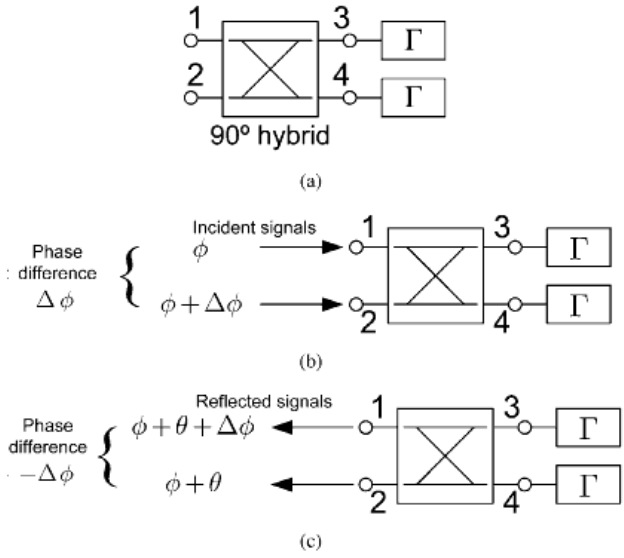

Fig. 2. (a) A terminated 90° hybrid with (b) and (c) to illustrate the phase difference reversal characteristics of the input signals and the reflected signals.

Figure 10. Passive VA array using hybrid couplers (reproduced from [7]).

A particularly attractive feature of this design is ease of implementation. Figure 11 shows 3-, 4- and 6-element retrodirective array circuits fabricated on 1.6 mm thick FR4 substrate using planar 90º microstrip branch line couplers. It is significant, too, that this retrodirective design does not require an even number of array elements as does the basic Van Atta design. The prototype array was fabricated with 2.9 GHz patch antenna elements on the same type substrate spaced 0.48λ. Measurements showed effective beam steering over a range of ±40º. This type of retrodirective antenna may be particularly attractive for the set-top application because of its simplicity.

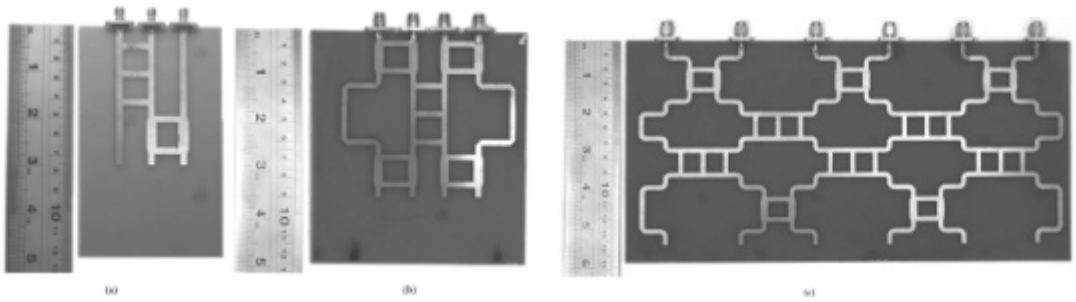

Figure 11. Passive 3/4/6-element VA array circuits as fabricated on PCB (reproduced from [7]).

Feed Systems for Van Atta Arrays. Two interesting and potentially useful VA array designs are described in [8]. Two planar VA arrays were designed and simulated using commercially available electromagnetic modeling software (ANSOFT HFSS). The first antenna (Figure 12, (a)/(b) left) used a microstrip line edge-fed patch at 7.58 GHz as the antenna element (overall dimensions ~13 x 30 mm). This element was arrayed in a two-dimensional 4x4 grid and evaluated for bandwidth and radiation characteristics. The second antenna comprised aperture-coupled microstrip patch antenna elements operating at 9 GHz also arrayed in a 4x4 2-D grid (Figure 12, (a)/(b) right). The second design was found to be superior because it could be more easily fabricated using both sides of the substrate and it provided better beam steering (±80º @ -10 dB). This simulation shows that very effective designs with small footprints are possible at very short wavelengths. This technology may be applicable directly or with modification or extension to set-top antenna requirements.

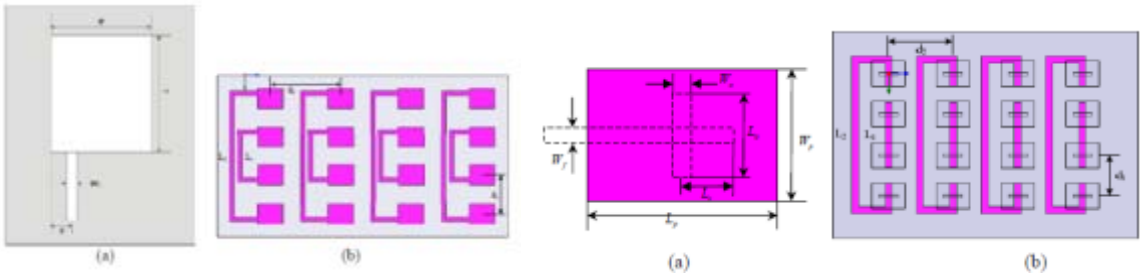

Figure 12. Planar edge-fed VA arrays (reproduced from [8]).

A Fractal VA Antenna Element. Dual-polarized Van Atta array antenna elements provide advantages over singly polarized elements, in particular increased transmit/receive (T/R) isolation on two-way links, multipath mitigation, and polarization diversity. While T/R isolation is not a consideration in the set-top television antenna, the other factors are. Additionally, size is a concern, and newer antenna element designs emphasize small size. An interesting fractal-based element is reported in [9] that may be useful for set-top VA arrays. Figure 13 shows the overall antenna in perspective and plan views. There are two square (40 mm) 1-mm thick material dielectric substrates, the top one being PTFE material ($\varepsilon_r = 2.65$), and the bottom FR3 ($\varepsilon_r = 4.4$). The First Minkowski fractal patch is shown on the right with the vertical and horizontal polarization feed ports marked V and H, respectively.

Details of the patch, which is designed to operate from ~4.4 to 5 GHz, appear in Figure 14. Its overall size is quite small at 21 x 16.8 mm, and it provides similar electrical performance to a square patch that is about 20% larger. Figure 15 shows details for the H-shaped slots that function as aperture coupling structures to excite the fractal patch. It is evident that, once dimensions are specified, the components are readily fabricated and easily assembled, which are very desirable characteristics for the set-top television antenna.

Computed data show that the antenna element exhibits H-polarization VSWR < 2 from ~4.82-5.17 GHz with 16.8% impedance bandwidth. The corresponding V-polarization figures are ~4.27-5.2 GHz, bandwidth 19.8%. Measured results confirm the modeling with bandwidth figures of 19.7% and 20.5%, respectively, which is good agreement. Directive gain is ~8 dB with >20 dB front-to-back ratio. These results clearly show that the microstrip Minkowski fractal antenna element is useful in VA arrays at microwave frequencies, and that it well may provide excellent performance for the higher television bands as well.

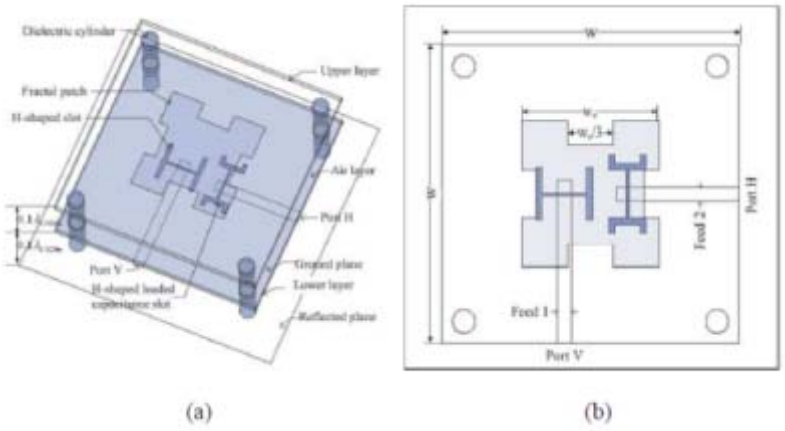

Figure 13. Proposed microstrip fractal antenna element (reproduced from [9]).

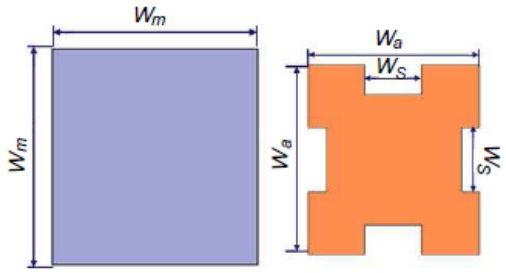

Figure 2. Structure of the conventional square patch and the Minkowski fractal patch (w_m = 21 mm, w_a = 16.8 mm).

Figure 14. Fractal patch (reproduced from [9]).

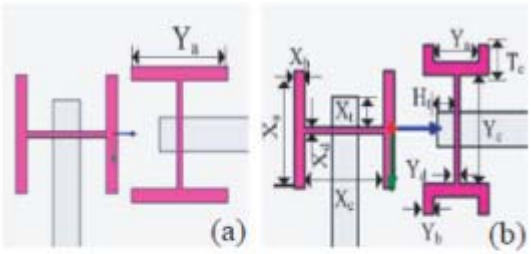

Figure 3. The aperture-coupled structures. (a) Double H-shaped slots, (b) H-shaped and H-shaped loaded capacitance slots.

Figure 15. Aperture details (reproduced from [9]).

Dual-Ring Antenna Elements. The dual-ring antenna element proposed in [10] is another novel device that can be used in retrodirective Van Atta arrays. Inner and outer concentric rings form an inherently two resonant-frequency structure. Each of the rings is parasitic with the other ring excited, and by properly choosing dimensions fairly widely separated resonance points can mark the upper and lower limits of a fairly flat response curve. The structure of the dual-ring element is shown in Figure 16. It is implemented in a two-layer microstrip structure with the rings deposited on the outer surface of an upper substrate underlain by a transmission line for feeding and a ground plane as shown. Element dimensions were optimized using HFSS modeling software, and the fabricated element tested. The measured bandwidth (-10 dB) was just under 34% (31-42.8 GHz), which agreed quite well with the predicted value of 34.3% (30-42 GHz). Between 30 and 40 GHz the average measured gain was ~4.5 dBi.

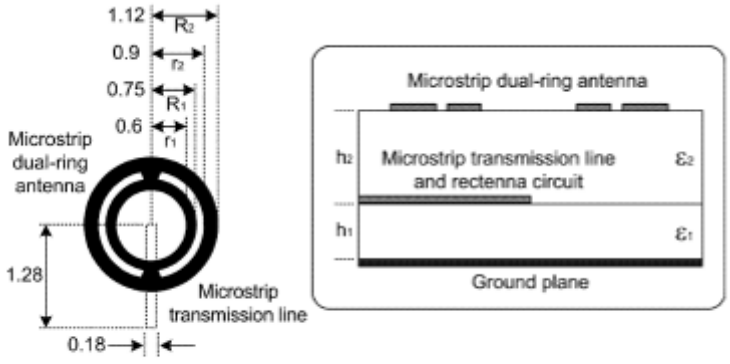

Fig. 3. Geometry of the broadband proximity-coupled dual-ring antenna and the two-layer dielectric structure. All dimensions are in mm.

Figure 16. Dual ring antenna element (reproduced from [10]).

The Van Atta array implemented using dual-ring elements was fabricated using 8 sub-arrays comprising 16 dual-ring elements in a 4 x 4 2-D grid. Figs. 17 and 18 show the sub-array and complete array structures, respectively. Measurements confirm the new array's predicted performance, which shows excellent beam steerability out to about

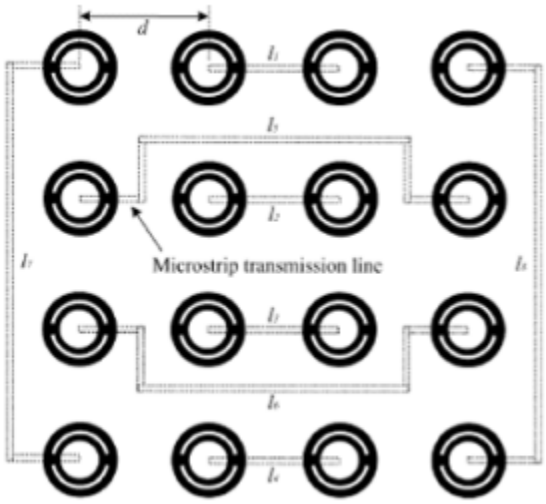

Fig. 5. Geometry of the 4x4 retrodirective sub-array where the dash-lines are the microstrip transmission line networks.

Figure 17. Dual ring element sub-array (reproduced from [10]).

±45º. The large array exhibits a very narrow main half-power beamwidth on the order of ~2º that is very stable between 32 and 49 GHz. Of course, this prototype antenna operates far beyond the highest television operating frequency, but its technology may be applicable to the TV set-top antenna with appropriate modifications.

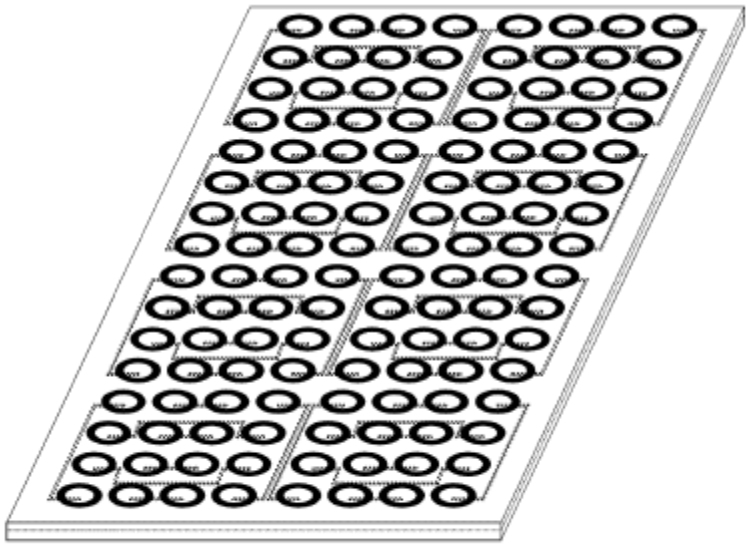

Fig. 6. Geometry of the 8x16 retrodirective array that consists of eight 4x4 sub-arrays.

Figure 18. Retrodirective array of dual ring elements (reproduced from [10]).

Conformal Retrodirective Arrays. In some cases it may be desirable to deploy a retrodirective antenna on a non-planar surface (the systems described above all utilize planar structures). Unfortunately the simpler Van Atta design cannot be used in such cases, so that Phase Conjugating systems are required instead. An interesting example of a genetically optimized, thinned PC array is described in [11]. The array's schematic representation appears in Figure 19, and its final physical configuration in Figure 20.

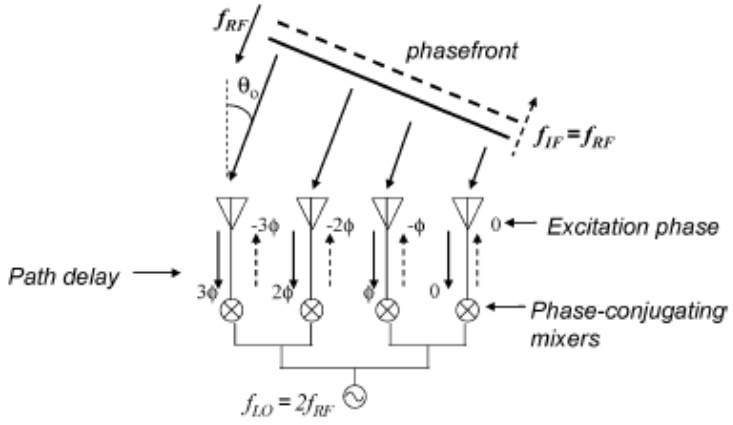

Fig. 1. A retrodirective array using heterodyne technique [5].

Figure 19. Conformal PC retrodirective array schematic diagram (reproduced from [11]).

In this example, the array was thinned to a total of 16 elements whose positions were optimized to achieve reasonable sidelobe levels over a scanning range of ~±120º. The shape and size of the optimized array aperture cannot be determined analytically, but a suitable optimization algorithm permits design of conformal PC arrays. This type of structure could be important in the set-top environment as a space-saving measure, but it would incur the additional expense and complexity associated with active PC arrays compared to the simpler passive Van Atta design.

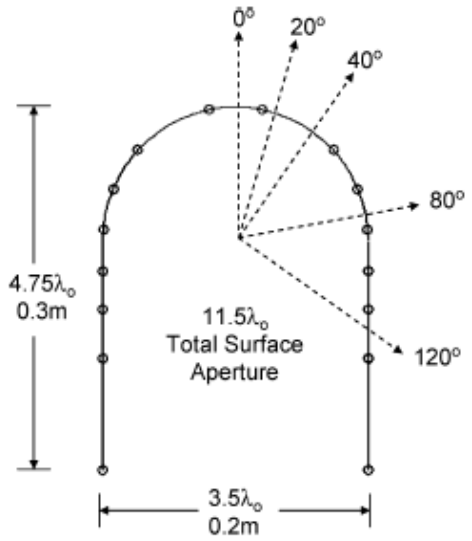

Fig. 2. The shape and the physical size of the aperture [4].

Figure 20. Conformal PC array structure (reproduced from [11]).

Near-Field Effects. Retrodirective antennas typically operate primarily in free space or near-free space environments, so that near field obstructions are not a factor in antenna performance. This typical environment is quite different from the set-top environment in which nearby objects and structures will definitely be in the antenna's near field and consequently a potential source of interference. Near field effects were studied in [12] in order to investigate these effects. Figure 21 shows the experimental antenna in an anechoic chamber and typical measured data. Three different near field scatterers were studied: (a) straight wire grid; (b) partially obscuring low loss MDF (medium density fiberboard); and (c) partially or totally obscuring concrete blocks.

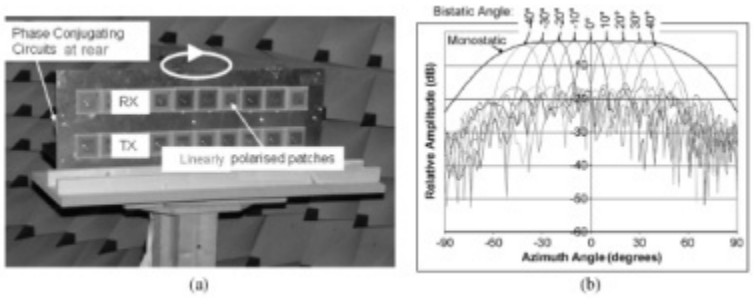

Figure 21. Measurement system for near field effects (reproduced from [12]).

Measurements were made at 2.4 GHz using a 10-element retrodirective array scanning ±40º with a half-power beamwidth of ~12º. The retrodirective antenna performed quite well in the presence of obstructions. For example, using gapped concrete block scatterers, it was able to fully compensate for phase distortions that reduced a basic reference antenna's gain by 11 dB. This study was the first to demonstrate that retrodirective antennas offer demonstrable performance advantages over other antennas in difficult propagation environments such as the interior of a building. This finding is an important reason for exploring retrodirective technology as it might be used in a television set-top antenna.

2.10.2 References

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3.0 Conclusions and Design Recommendations

Upon consideration of Table 1-1 at the end of Section 1 of this report we have concluded the following:

- There have been significant and potentially useful antenna design methods and actual hardware developed over the past 15 years that will improve the performance of low-profile, compact indoor/set-top DTV antennas.
- In the area of electro-magnetic computational and design optimization methods Genetic Algorithms (GA) and the Central Force Optimizer (CFO) have proven themselves as powerful tools in the design of broadband, compact antennas.
- Using the above, we have included an example of an advanced antenna technology, specifically the Fragmented Antenna as part of this project. We submit that it would be nearly impossible for a human to replicate its design and performance.
- The Non-Foster active broadband matching technology and its required semiconductors have matured to the point where they could be incorporated into a commercial indoor/set-top design to provide a matched antenna where it is considered as being electrically small. In the example shown below it would be used to provide acceptable performance with the Fragmented dipole within the 54-88 MHz band. Other element geometries are also possible.
- Many of the other advancements listed in Table 1-1, specifically 2.4 through 2.6, require a CE-909-A type interface with the DTV receiver, and while potentially useful if and when manufacturers start adding this feature to their products they offer no practical improvement for the near term.
- The technologies designated as 2.7 through 2.10 while interesting are either too embryonic (2.7 & 2.8), too complicated and too large (2.10) or not well vetted (2.9) to be considered in the near term.

Considering the above it is recommended that the NAB give consideration to further developing a computer generated/optimized broadband dipole antenna element, such as a Fragmented Planar Dipole with/without a Non-Foster Matching circuit in order to demonstrate that it is feasible to design a VHF/UHF antenna system that is both simple, have positive gain, be of reasonable size and weight and not require a CE-909-A interface with the DTV receiver.

As an example, a fragmented planar antenna was optimized for 54-698 MHz to cover all three DTV television bands. The antenna has dimension of 13” x 13” across the VHF and UHF bands, a block diagram of the proposed antenna system is shown in Figure 1. The computed directivity, after optimization, along the horizon is good at all DTV frequencies as shown in Figure 2. The antenna shows a good match at UHF frequencies, but as shown in Figures 4 and 5 some matching, using a simple L-C passive network of SMT components and Non-Foster Matching circuitry at the feed-point would be necessary for good performance in the high and low VHF bands respectively.

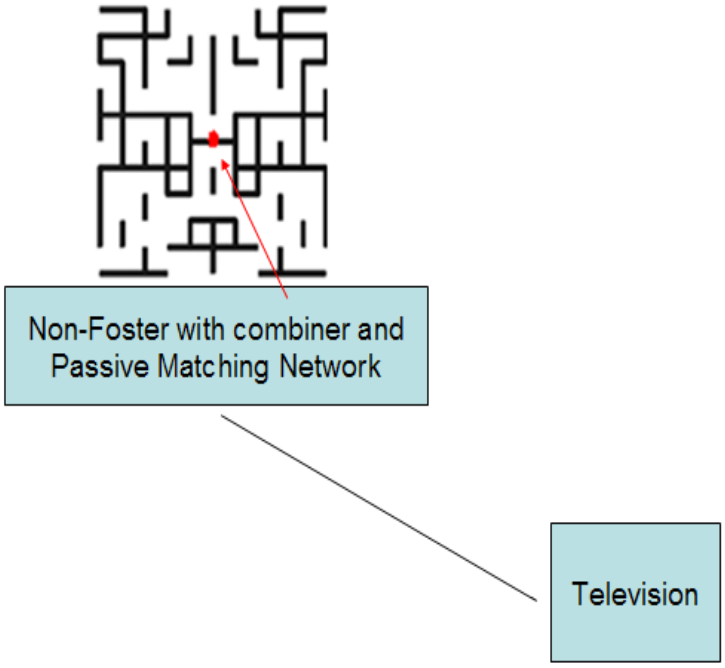

Figure 1. Fragmented Antenna Example, 13”x13”

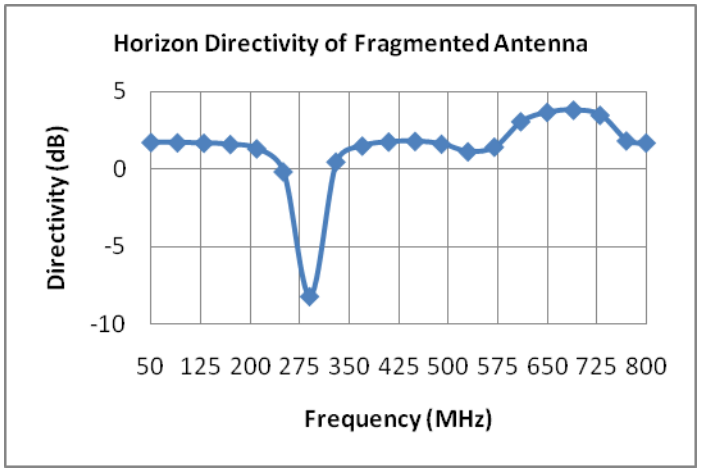

Figure 2. Directivity Along Horizon for Fragmented Antenna

The pattern of the antenna as currently designed is a “figure eight” dipole pattern at all frequencies, as shown in Figure 3. Figure 6 indicates that without any matching, a very low VSWR across the UHF 470-698MHz band.

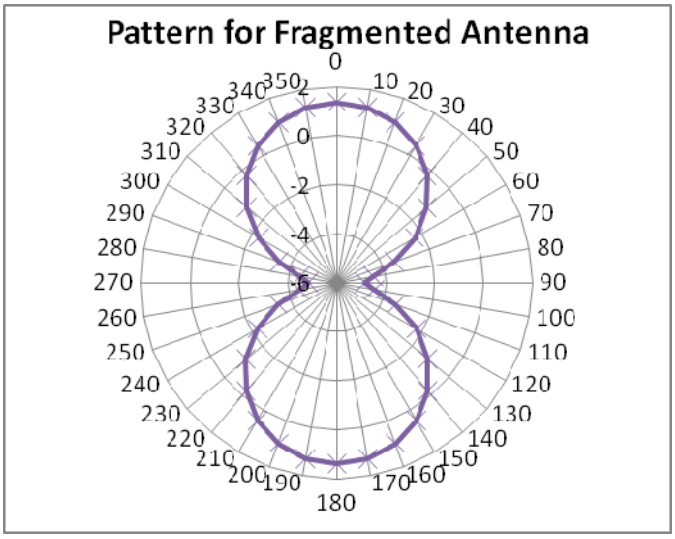

Figure 3. Azimuthal Pattern of Fragmented Antenna

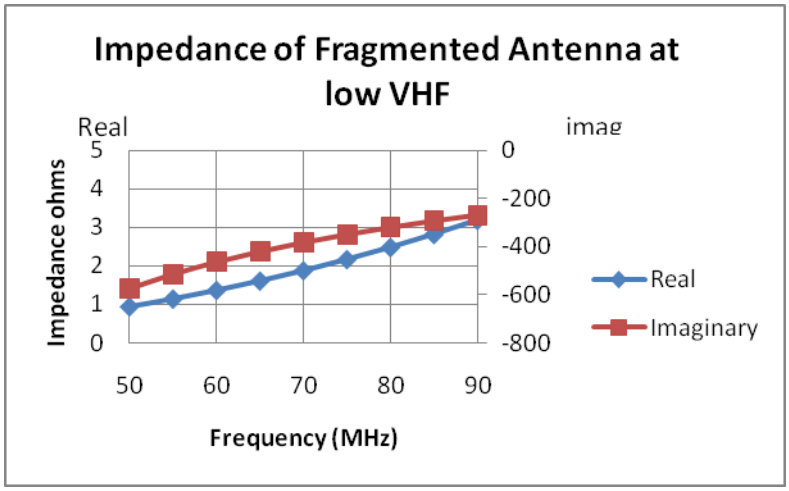

Figure 4. Impedance of Fragmented Antenna at low VHF

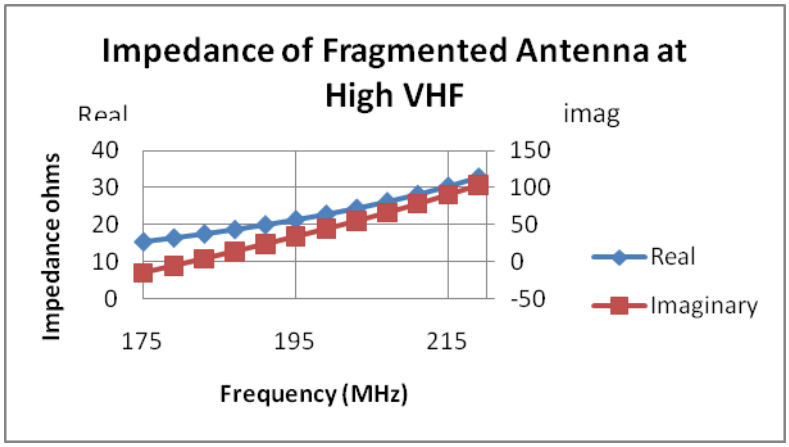

Figure 5. Impedance of Fragmented Antenna at high VHF

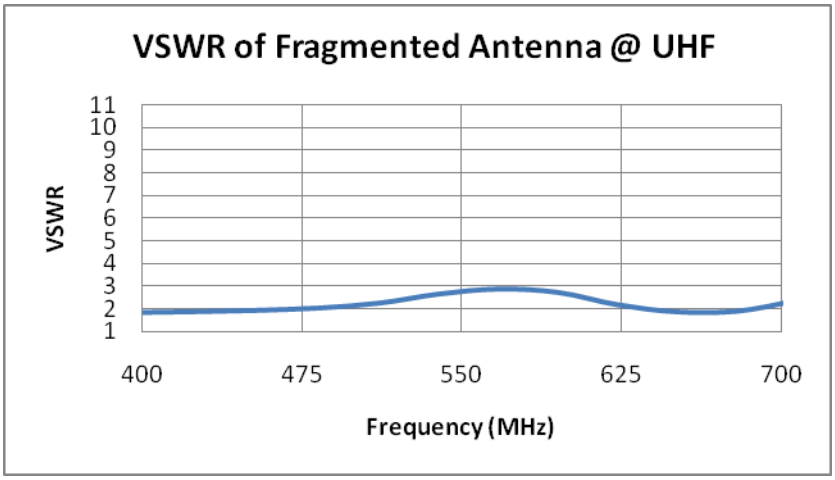

Figure 6. VSWR of Fragmented Antenna at UHF

A Microstrip Patch Antenna Manufactured with Flexible Graphene-Based Conducting Material

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***Abstract*—In this paper, a unique process for fabricating microstrip patch antennas with flexible graphene-based conductors is presented. In particular, this manufacturing process uses a commercially available micro-cutter to cut the outline of the patch antenna from the flexible graphene-based conductors, and then this piece is attached to a grounded FR4 substrate using adhesive to create a unique printed antenna. The design was modeled using a commercial simulator, and a prototype was fabricated and measured. Overall, it was shown that the S-parameter simulations agreed fairly well with measurements, and that this manufacturing process has the potential to develop more complicated designs, such as meander-line dipoles for example, that are difficult to cut-out manually.**

***Index Terms*—Patch antenna and graphene-based conductors.**

I. INTRODUCTION

Modern wireless communication systems are being required to operate in evermore complicated environments. This is because these systems are being applied to problems that (1) involve surfaces that change shape with time, (2) include wearable networks and/or (3) are subjected extreme environmental temperature/pressure changes [1]. Because of the wireless nature of these systems, the antenna is a major part of the design and conformal antennas have the potential to overcome some of the aforementioned difficulties. However, one drawback of using a traditional conformal antenna can be the copper conductors. This is because of the weight and the potential for copper failure due to repetitive bending deformations, which was noticed in the work reported in [1]. To mitigate some of these issues, a flexible graphene-based conducting material [2] is being explored as a possibility of replacing the etched copper in a conformal antenna. A picture of the material is shown in Fig. 1(a). However, an efficient manufacturing process for fabricating complicated antenna designs based on this material is required. Thus, the objective of this paper is to present a suitable manufacturing process and introduce the graphene-based microstrip patch antenna shown in Fig. 1(b). This provided a standard design that has been well understood and was marginally complicated.

An initial investigation on graphene patch antennas at microwave frequencies was reported in [3] and later discussed in [4]. The authors of these papers summarized the benefits of graphene as a mono-atomic structure, whereas the presented work builds on these ideas and explores the potentials of using graphene-based conductors that are commercially available and much thicker than a single layer for printed antenna purposes.

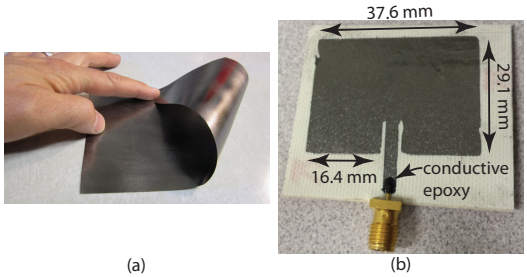

Fig. 1. Photograph of the commercially available flexible graphene-based conducting material and (b) a photograph of the prototype (with dimensions) of the fabricated graphene-based patch antenna.

II. THE MANUFACTURING PROCESS

Several steps are required to prepare the graphene-based conductor material in Fig. 1(a) for manufacturing. Initially, the temporary spray adhesive (Sulky KK 2000 [6]) shown in Fig. 3 is applied to the top surface of a 100 μm thick sheet of paper and bonded to the bottom surface of the graphene-based conducting material (Fig. 2). Next, the same spray adhesive is applied to the bottom surface of a 100 μm thick transparency film and bonded to the top surface of the graphene-based conducting material (also shown in Fig. 2). As a result, the three layers are bonded together for manufacturing, giving the 3-layer structure shown in Fig. 2. The 3-layer structure is then placed on the adhesive cutting mat shown in Fig. 3. The graphene-based conducting layer is now ready for cutting out the shape of the microstrip patch. It should be noted that the three layers were pressed together by hand and the temporary adhesive was cured at room temperature for 2 - 3 minutes before cutting. This allowed for the separation of the three layer after the cutting process was completed. Finally, the layout of the patch is defined in the software included with the micro-cutter, which is discussed in the next section.

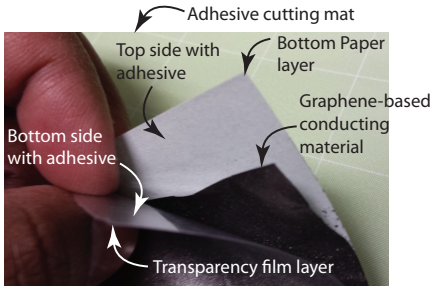

Fig. 2. Photograph of the flexible graphene-based conductors placed in between different layers to fabricate the antenna.

Fig. 3. Photograph of the commercial micro-cutter [5] and the temporary adhesive spray [6].

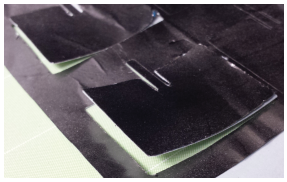

Fig. 4. Fabrication step to separate the flexible graphene-based conducting layer of the patch from the protected layers.

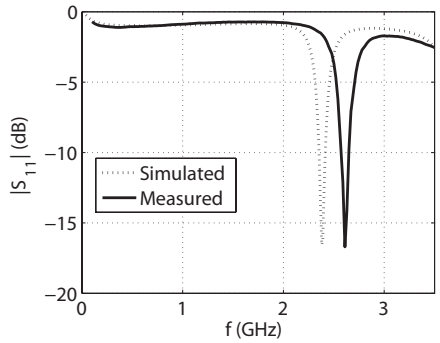

Fig. 5. Measured and simulated $|S_{11}|$ values of the prototype graphene-based antenna.

III. Design and Evaluation of the Prototype Microstrip Patch Antenna

To develop the patch antenna in Fig. 1, the conductive properties of the graphene-based material [2] were determined by measuring the S-parameters of a known printed microstrip transmission line that used the graphene-based conductors instead of copper. The conductivity was determined to be σ = 1.94×10^5 S/m for a thickness of 25 μm. Next, this thickness and conductivity were used in the commercial simulation software ADS [7] to determine the geometry of the patch antenna shown in Fig. 1(b). The simulated S-parameters are shown in Fig. 5. The resonant frequency was predicted to be 2.4 GHz and the grounded FR4 substrate was 1.5 mm thick with ε_r = 4.5 (the bottom layer of the FR4 was 0.5 oz copper). Then, this geometry was drawn in the software included with the micro-cutter and used to cut out the graphene-based conducting layers shown in Fig. 4. As a final step, the graphene-based conducting patch was removed from the 3-layer structure and attached to the FR4 substrate. The result was the prototype shown in Fig. 1(b). The center conductor of the SMA connector was attached to the patch with conductive epoxy. Next, the $|S_{11}|$ values of the prototype were measured in an anechoic chamber and are shown in Fig. 5. The resonant frequency was 2.6 GHz.

Several comments can be made about the results in Figs. 1(b) and 5. Closer observation of the edges in Fig. 1(b) shows that the micro-cutter blade tends to wobble slightly during cutting, which may be the cause of the differences in the S-parameters in Fig. 5. This behaviour could be due to the blade depth setting and further exploration is planned in the near future. On the other hand, the results in Fig. 5 show that a new patch antenna can be developed using this flexible graphene-based material, which may be difficult to otherwise manufacture with the existing PCB techniques.

IV. Conclusion

An initial investigation of a new patch antenna that uses flexible graphene-based conductive material was introduced in this paper. Additionally, a unique process of manufacturing these patch antennas with the graphene-based material was presented. The design was simulated in ADS using extracted material properties, and a prototype was fabricated and tested. Overall, fair agreement between the $|S_{11}|$ values was observed and it was demonstrated that moderately complex geometries could be manufactured using a simple and cost-efficient commercially available micro-cutter.

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BUILD A SHORTWAVE ANTENNA

"The Search For The Perfect Shortwave Antenna"
by N4UJW Webmaster Hamuniverse.com

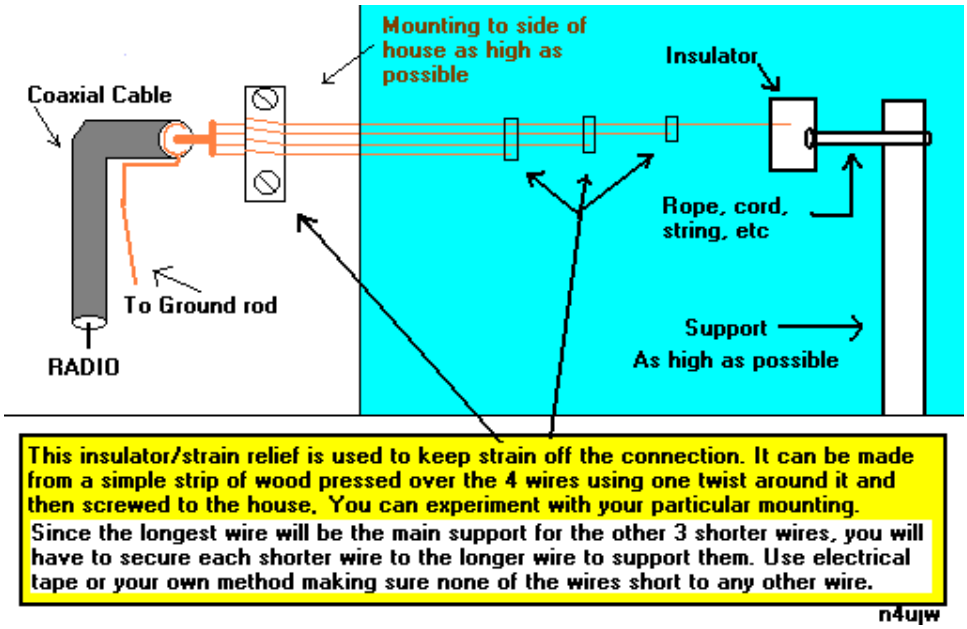
THE MULTIBAND LONG WIRE SHORTWAVE ANTENNA

A Much Better But More Complicated Antenna

This antenna is end supported and designed to receive the major shortwave bands between 90 meters and 16 meters. It uses only 4 wires and a unique antenna property called harmonics to get 8 bands using only 4 wires! Again, it is a compromise but an excellent performer....the perfect antenna does not exist. We "Hams" are working on it constantly!

After construction, this shortwave antenna should be stretched out in a straight line as high as possible as in the long wire antenna above, and about 140 feet straight out from the house! Don't fret! If you cant', you can't. Utilize your existing space. More supports may be required for a zig zag layout but performance may suffer a bit. Don't worry, it will certainly outperform that built in poor excuse for an antenna!

It consists of 4 separated insulated wires, (measurements below), all connected (soldered) on one end, leaving the opposite end unconnected and insulated at the support. If you do not know how to solder, then scrap all the coating from the wire down to bare copper and tie the ends together using several knots. You really should learn to solder though!. This will make for a more permanent and much better electrical connection.
The soldered end must be between an insulator and the radio for mechanical strength.
You don't want much stress on the soldered connection other than the coax leading to the radio. The end that has all wires connected should be soldered to the center wire of a suitable length of 50 - 75 ohm coaxial cable leading to the short wave radio with a suitable connection. A ground wire is soldered to the shield only of the coax at the same end that you soldered all the wires together and attached to a ground rod driven into the ground near the house. Seal and tape all outdoor connections from the weather. This antenna is called an end fed half wave antenna.
See picture, formula and wire measurements for bands below:
(The lengths are not extremely critical, but try to get them as close as possible.)



Note: In the instruction box above, the last sentence refers to the long portion of the wires, not at the connection point to the coax feed line to the receiver. All wires are connected together at the connector center conductor wire!

(frequencies shown below are approximate shortwave band centers):

Wire 1 (LONGEST WIRE) 3.25 MHz (90 meter band) 09.75 MHz (31 meter band 3rd harmonic)
468 ÷ 3.25 = 144' 0"

Wire 2 3.95 MHz (75 meter band) 11.85 MHz (25 meter band 3rd harmonic)
468 ÷ 3.95 = 118' 6"

Wire 3 5.10 MHz (60 meter band) 15.30 MHz (19 meter band 3rd harmonic)
468 ÷ 5.10 = 91' 9"

Wire 4 (SHORTEST WIRE) 5.90 MHz (49 meter band) 17.70 MHz (16 meter band 3rd harmonic)
468 ÷ 5.90 = 79' 3"
The number 468 divided by the frequency above is the formula for calculating a half wave antenna length used all the time by Amateur radio operators in building many different kinds of antennas.
You'll need about 435 feet of wire for this antenna plus appropriate length of coaxial cable.

The wires are spread 3-4 inches apart, held in place with simple non-conductive spacers.
Just cut a few pairs of the acrylic, Plexiglas, plastic strips or other non-conductive material that will not be damaged by moisture long enough to attach the wires keeping the spacing about 3 to 4 inches or further if you want.

OUTDOOR CENTER FED MULTIBAND (FAN) DIPOLE SHORT WAVE ANTENNA

NOTE: For use with the higher quality table model communications receivers that have standard antenna connectors capable of using direct coaxial cable connectors.

This antenna type is used by many Ham Radio Operators worldwide and is very popular but the lengths for the Ham bands are entirely different.
The entire length of the antenna is about the same as the one above and the coaxial cable is connected in the center of the span with the center conductor connected to one side of the antenna and the shield connected to the other side then at the other end, to the receiver.
The formula used for this antenna is the same as the Multiband Long Wire above:

468 / by frequency in mhz = total length in feet. This resulting length is cut in half!

One antenna per band stacked.

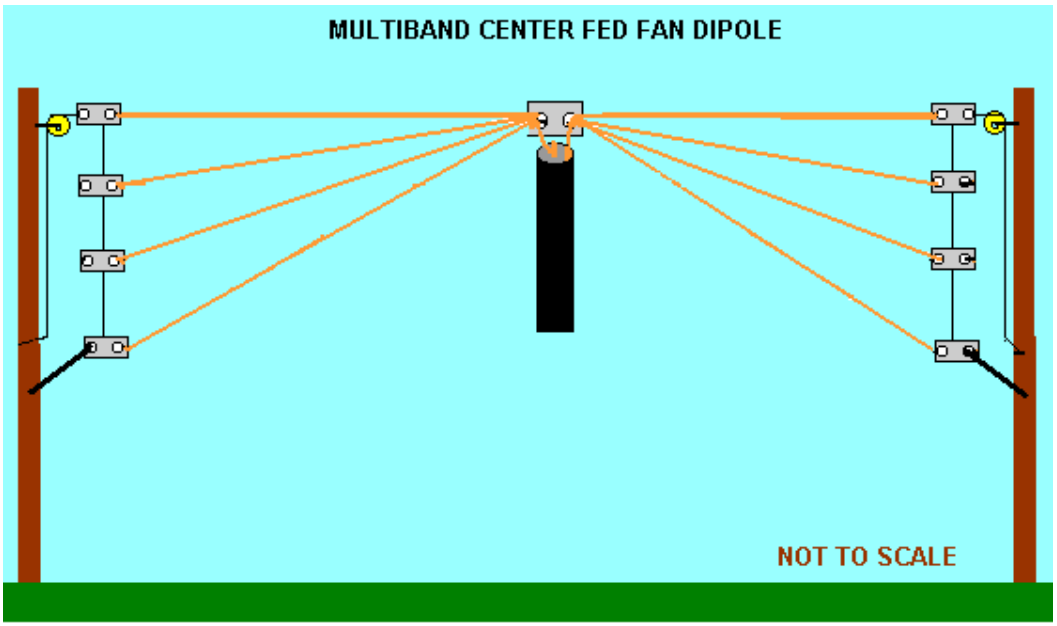
It is somewhat more complicated in construction due to the center connection and requires support in three places....each end plus the center. The preferred method for using this antenna is drawn in the picture below with the wires "fanned" apart with at least a foot of separation between the ends.
All of the wire elements can be close spaced but some interaction will occur. Insulated wire is best so the individual wires do not connect on the longer lengths of the antenna.

Choose the antenna of your choice depending on your constructions skill and needs.
Either way, they both will be much better than the little telescoping antenna that comes with most portable receivers.

The center fed multiband dipole antenna (drawn below) consists of 2 separate sections of 4 wires on each side of the center connection at

the support consisting of 4 wires connected to the center conductor of the coax and the other 4 connected to the shield.

In this arrangement, one half of the antenna is feeding the center conductor and the other half is feeding the shield. Each side must be insulated....not connected....to the other side. The other end of the coax is connected to the radio with the appropriate connector.
Use lengths in the above multiband antenna with total length split in half using the formula....half on one side and half on the other for each wire length per band. The coax can be anything from 50 ohm to 75 ohm.
Not critical on receive!



Shown supported between two wooden poles. Gray rectangles are insulators.
Yellow circles are pulleys.
Orange lines are dipole elements.
Heavy black vertical line in center is coax leading to radio.
Note that one side of antenna is connected to center of coax, shield to the other side.

The final assembled antenna can be installed with the center section higher than the ends, making it look like an inverted V, like this /\ .
Make the angle of the V about 90 degrees or more.
Or it can be horizontal to the earth or anywhere in between.
The inverted V configuration is more omni-directional, (all directions), than the horizontal method which tends to receive best, broadside to the wire. Less real estate is required for the inverted V method. Center supporting also has less tension on the antenna so smaller wire size may be used to save money.

Choose the antenna of your choice above depending on your constructions skill and needs. Either way, they both will be better than the little telescoping antenna that comes with most portable receivers. Enjoy!

INDOOR MULTBAND SHORTWAVE ANTENNAS

Attic Antennas

To begin with building and installing an attic antenna that helps your reception, you need to take stock of your attic's measurements, particularly the length of the attic at it's longest distance that you have easy access and your radio's location.
One of the more common house sizes is about 50 to 60 feet long and about 25 to 30 feet wide at the ground level. Your house or home may be entirely different. The accessible attic space usually is much less than this. You will have to really compromise with an attic antenna as far as the band coverage is concerned for a short wave antenna to perform adequately. Use the dimensions of your attic and compare them with the lengths of the long wire and dipole type antennas in this article above and choose the one that you can "fit" into the attic. You may not be able to use lengths for all the bands, but again, no matter what length your end result is, it will certainly outperform that little pip squeak of a poor excuse for an antenna that came with the radio! Just utilize the space that you have and don't worry about the length. Just use as much wire as you can and forget about that "perfect antenna". It still does not exist up to this point in this article! Hams are still working worldwide on it!

The best place to mount or attach the antenna is against the peak or highest part of the roof thereby keeping it away from ductwork, AC and heater systems, telephone and all the other metallic environment that exists in most attics. Once you have the location selected, then build the antenna while keeping in mind that the coax or wire will have to get to the radio. If you're working up on the roof, get a helper to assist, an adult, not children! Be careful on those ladders!
You can push most small coaxial cable under the space where the carpet and wall come together and wire should be no problem, then to the nearest closet, up the wall and into the attic. You can work from the attic down or radio up....your choice. Lots of variables here too so you will have to choose your own route and method of installation. If you have to drill into a wall to feed the wire, use caution and don't drill into electrical wires! It may be the last time you do!

IN ROOM ANTENNAS!

In those cases where you can't put an antenna outside or up in the attic, then you can install it in the same room with the radio! They won't be as effective as those up in the attic or outside but will still get more signal to your radio which is what you want.
Simply use your own method to attach a random length wire, up next to the ceiling against the walls...around all sides of the room if possible. One other choice is to push a random wire between the carpet and the baseboard around the walls of the room. You will be surprised at the difference compared to that telescoping antenna that came with your radio. Just attach the antenna to the telescoping rod...don't forget to remove the insulation on the wire at the attachment point!

Check out the Ham Radio Antenna Projects here!

A Metamaterial-inspired Approach to RF Energy Harvesting

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Abstract

We demonstrate an RF energy harvesting rectenna design based on a metamaterial perfect absorber (MPA). With the embedded Schottky diodes, the rectenna converts captured RF waves to DC power. The Fabry-Perot (FP) cavity resonance of the MPA greatly improves the amount of energy captured. Furthermore, the FP resonance exhibits a high Q-factor and significantly increases the voltage across the Schottky diodes, which improves the rectification efficiency, particularly at low power. This leads to factor of 16 improvement of RF-DC conversion efficiency at ambient intensity level.

Introduction

Low-power RF energy harvesting has become a field of interest due to the need to acquire power in situations where the use of wires and/or batteries is impractical such as in the cases of expansive sensor networks and structural health monitoring. The recent proliferation of RF signals due to cell phones, Wi-Fi networks, and GPS has produced readily available ambient power sources (although of low power density) for scavenging energy. Historically there has been difficulty with creating omnidirectional electrically small energy harvesting antennas at RF frequencies that efficiently collect energy[1]. There is further difficulty with efficient AC to DC rectification in cases of low input power[2]. In this paper we demonstrate the potential of metamaterial-based energy harvesting antennas to address both of these issues[3].

Research in metamaterials is a relatively new and unexplored field of study with objectives for constructing materials with customized optical, mechanical, or electrical parameters that would otherwise be difficult or impossible to obtain with ordinary materials[4,5]. Metamaterials are constructed by building a 2-D or 3-D array of structures within a substrate. The choice of the structures used in the array is the primary determinant of the properties of the bulk metamaterial. Thus the bulk properties of the metamaterial can be tuned by altering the shape, spacing, size, etc. of the structures that the metamaterial is composed of. Provided that the size of the structures is sufficiently smaller than the wavelength of the incident electromagnetic waves, the waves interact with the structures in a fashion analogous to the way that optical light or x-rays interact with the atoms and molecules of ordinary materials. There are a multitude of potential applications for metamaterials such as the construction of negative index materials[6,7], cloaking[8], perfect lenses[9], and energy harvesting[10] (which will be the focus of this paper).

A big limitation of metamaterials is that they will typically only function as intended within a limited bandwidth. The structures used to make a metamaterial typically resonate such that the metamaterial exhibits the desired properties only at its resonance frequencies. Thus the metamaterial

functionality is often confined to a relatively small bandwidth surrounding each resonance peak, the

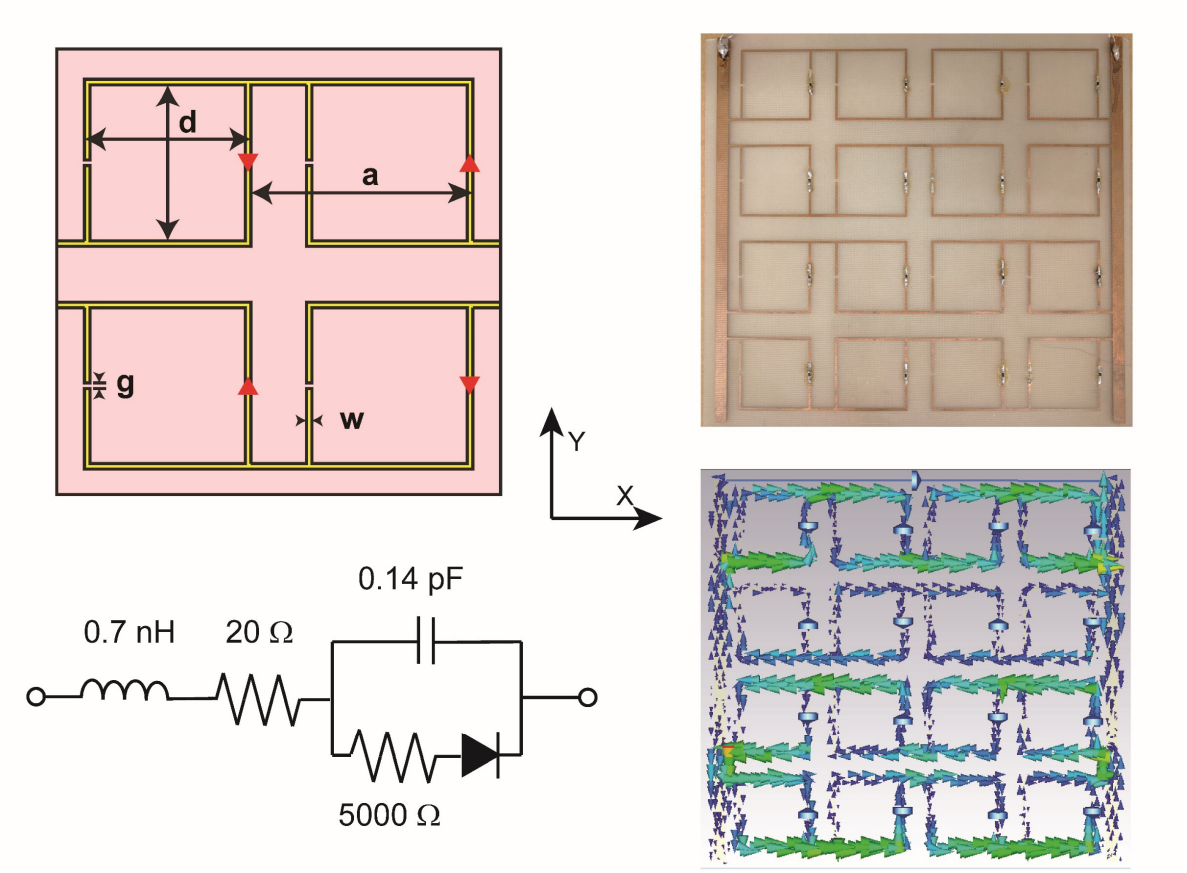

Figure 1 - Rectenna sample design with the dimensions of the unit cell and the equivalent circuit model used for the Schottky diodes in simulations. Diodes are indicated in the unit cell by red triangles pointed in the direction of current flow. d = 30 mm, a = 20 mm, g = 1mm, w = 1mm, t = 0.8128 mm (thickness of FR4 substrate), s = 0.0178 mm (thickness of copper ring layer). The incident wave is perpendicular to the plane of the rectenna and polarized in the x-direction. The surface current modes activated at the resonance frequency (908 MHz) are shown on the sample.

performance of which depends on the Q-factor of the resonance.

An antenna constructed out of an appropriately designed metamaterial that incorporates a means of rectification can operate as an efficient energy harvesting system that can be used to convert power captured from RF sources (or possibly other regions from the electromagnetic spectrum) to useful DC power [11,12]. Optimal RF energy harvesting occurs when the metamaterial is constructed such that a high Q-factor resonance occurs at the same frequency as the external radiation source. For ambient RF energy harvesting, these frequencies will usually be found in the ISM band (particularly those from 800 MHz to 6 GHz) from sources such as cell phone signals, GPS, and WI-FI. References [13-15] provide measurements of power density levels that one might expect to find in urban environments. The greatest power density measured was 6.4 $\mu W/cm^2$, but average values typically ranged from 1-50 nW/cm^2 depending on the frequency, time of day, and proximity to RF sources.

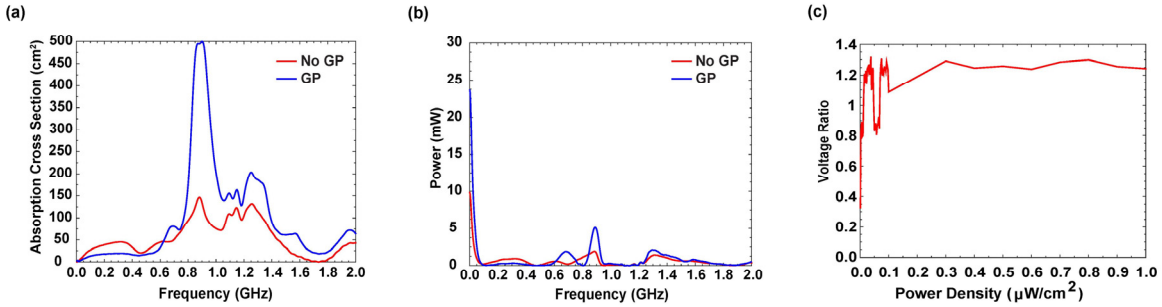

Figure 2 - (a) Comparison of simulated rectenna absorption with and without the ground plane. The absorption peak narrows and increases, indicating an increase in the Q-value of the resonance. (b) Simulated power delivered to the load from a Gaussian pulse. The DC component of the power delivered to the load is the largest component, suggesting good rectification efficiency. The DC component is also strongly enhanced by the ground plane. (c) Oscilloscope measurement of the DC voltage across one of the Schottky diodes. The presence of the ground plane enhances the voltage, which makes it easier for the diode to activate (particularly at low power) and thus improves energy harvesting efficiency.

Reference [16] gives a nice summary of past attempts at developing high efficiency low-power energy harvesting systems for various power densities and frequencies. As mentioned by the authors, the results between groups are not directly comparable due to differing frequencies, definitions of efficiency, and power densities. Nevertheless, they provide a reference point for evaluating the results presented by this paper. Other efficiency benchmarks are presented by references[17,18]. The results of our measurements on the energy harvesting efficiency of the metamaterial sample described in figure 1 are presented in this paper. Upon examination of the results presented by other groups, we believe these results to be the most efficient over the power densities and frequencies investigated.

Results

The absorption cross section obtained by simulation of the rectenna (figure 2a) predicts a peak energy harvesting frequency around 0.8 GHz. Adding the ground plane causes the peak to become somewhat narrower, but much larger (an increase in Q-value). As seen in figure 2b, the power delivered to the load is primarily from DC, while power from AC signals is much smaller. When the ground plane is incorporated, the DC power is much higher. A rough oscilloscope measurement of the voltage across one of the diodes shows the voltage-enhancing effect produced by inclusion of the ground plane (figure 2c).

Efficiency measurements for relatively high power densities (figure 3) indicate that the maximum energy harvesting efficiency occurs at 0.9 GHz, which is comparable to the frequency predicted by the absorption cross section. The power density range is well above what would typically be expected to be available from ambient RF signals and thus would only be found within close proximity (<100m) to a strong RF source such as a cell phone tower or right next to a weak RF source. In general, the efficiency of the sample increases with incident power density. However, when the ground plane is present, there is a point where the efficiency begins to diminish with increased power density (the red curve in figure 3f). When the ground plane is placed at the optimum distance (4cm), the energy harvesting efficiency improves considerably (compare figures 3a-3c with 3d-3f) and goes well over 100% at times. This indicates that the ground plane has caused the effective area of the rectenna to become larger than its geometric area.

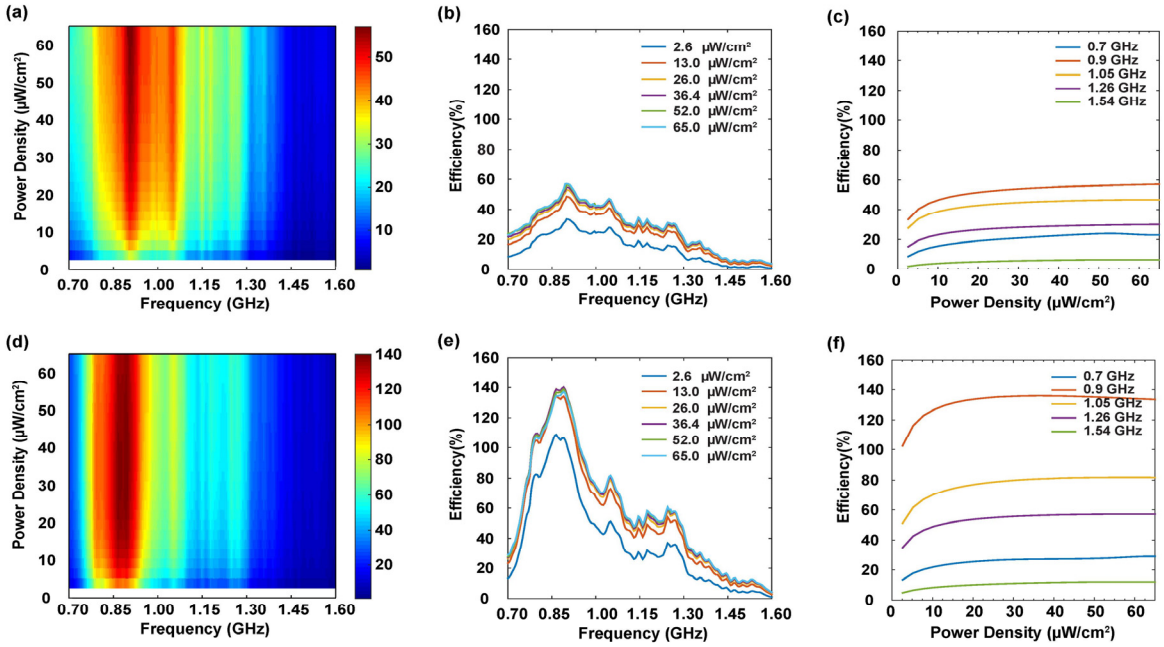

Figure 3 - Experimental measurements of energy harvesting efficiency as a function of frequency and power density. (a)-(c) are measurements without the ground plane. (d) - (f) are measurements with the ground plane.

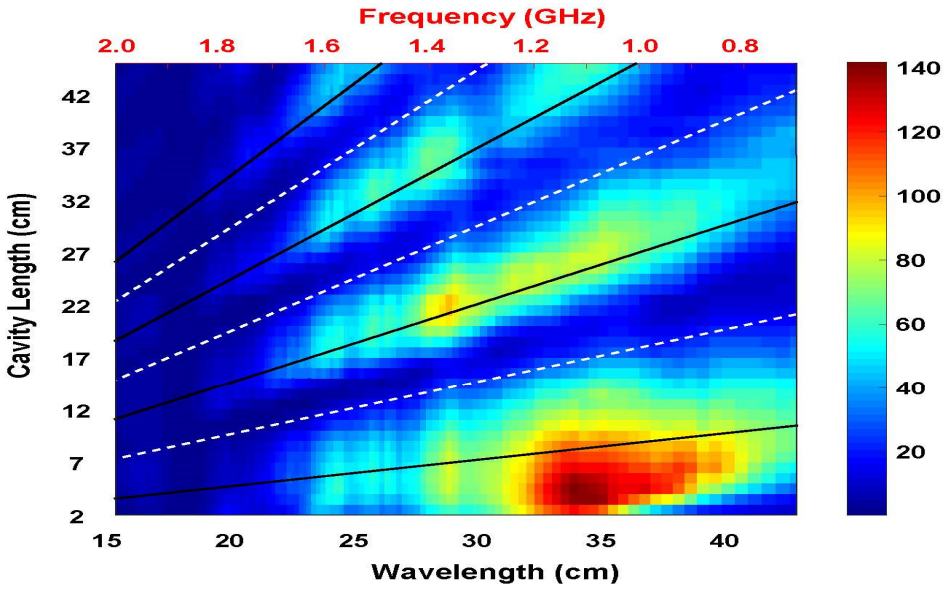

Figure 4 - Measurement of the effect of Fabry-Perot cavity length on energy harvesting efficiency at fixed power density (30 μW/cm²). Solid Lines indicate the theoretical prediction where maximum energy harvesting efficiency would occur for a vacuum-filled cavity with half-open boundary conditions. Dashed lines indicate where energy harvesting maxima would occur for a vacuum-filled cavity with closed boundary conditions. The metamaterial perfect absorber rectenna behaves similarly to a half-open FP cavity. The shortest observed resonance length (λ/8) is less than predicted (λ/4), because the effective index of refraction of the metamaterial surface is larger than one.

Figure 4 demonstrates the efficiency-enhancing effect of the Fabry-Perot cavity created by the rectenna and the ground plane. Examining the relationship between the efficiency, ground plane distance, and RF wavelength reveals standing wave patterns resembling that of a resonating cavity with one side closed (ground plane side) and the other side open (rectenna side). In the figure, the standing wave pattern is compared to the theoretical predictions of vacuum-filled Fabry-Perot cavities with half open (solid black lines) and closed (dashed white lines) boundary conditions.

Figures 5a and 5b show the energy harvesting efficiency measurements at power densities that are comparable to what can be found from ambient sources (up to 1.0 $\mu W/cm^2$). The power density levels presented were chosen to correspond to what can be found within relatively short distances from a commercially available high-power (100 mW) omnidirectional WIFI antenna with a maximum gain of 3/2. Now the efficiency is clearly much lower than it was for the high-power measurements. However, the presence of the ground plane increases the efficiency by a large factor (up to 16) in this case. A plot of the ratio of the efficiency with the ground plane to the efficiency without the ground plane, which will be referred to as the enhancement factor, is shown in figure 5c. The presence of the ground plane produces the largest enhancement factor at very low powers (about 0.01 $\mu W/cm^2$, which is at a typical ambient level), but unfortunately the overall efficiency at this level is still very low (only about 1%). At the lowest power density levels the efficiency is no longer enhanced much.

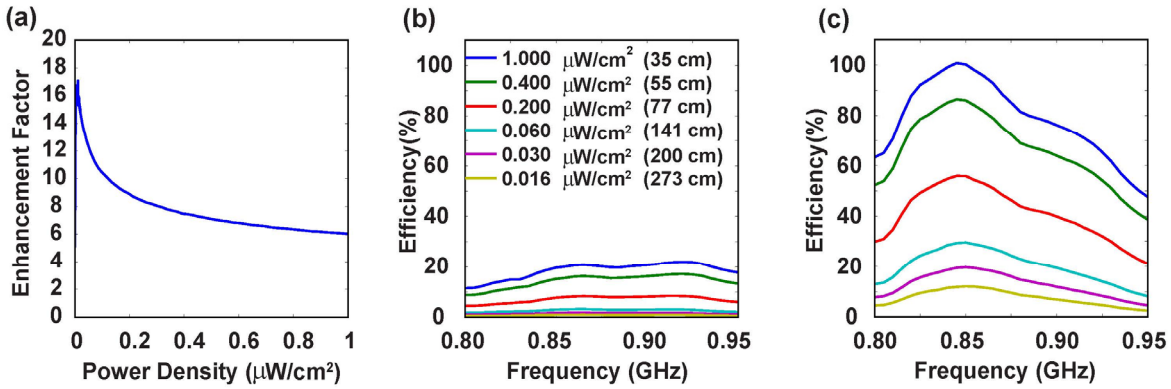

Figure 5 - Measurements corresponding to power densities that would be available at the specified distances from a 100 mW source possessing a transmitting antenna with gain of 3/2. (a) No ground plane. (b) Ground plane present. Same legend as 5a. (c) Enhancement factor; defined as the ratio of power harvested with the ground plane to power harvested without the ground plane. Measured at 0.85 MHz. The peak frequency is slightly different than before because of slight differences in the angle of incidence of the rectenna between measurements.

Discussion

The data indicates several advantages to basing an energy harvesting antenna off of a metamaterial perfect absorber. First, building the diodes into the metamaterial surface results in highly-efficient rectification at the resonance frequency as determined in both the simulated results (figure 2) and the experimental results (figure 3). Alignment of the FP resonance with the natural resonance of the metamaterial surface leads to the best results. Second, the tunability of the metamaterial's dielectric constant allows the rectenna to be designed such that it is impedance-matched to free space, and hence has a reflection coefficient near zero. This improves the absorption and has the added advantage of allowing for a smaller rectenna. Since the surface behaves closely to an open-boundary for the FP cavity,

the resonance length is shorter than for a closed boundary cavity (λ/4 compared to λ/2). The measured FP resonance peak occurred at a wavelength of λ/8 instead of λ/4 (figure 4), because the index of refraction of the material is not quite equal to 1. Third, as evidenced by the measurement in figure 2c, the conducting ground plane increases the voltage across each of the diodes in the rectenna. The effect gets stronger at low power (figure 5c) since it can provide enough of a voltage boost to activate the diodes when the diode threshold voltage would not ordinarily be attained. Once the power gets too low, then the diode threshold can not be met even with the ground plane present. If the power is too high (well above ambient levels), the diodes reach their breakdown voltage and efficiency decreases as shown in figure 3f. Fourth, the ground plane increases the directivity of the antenna (and thus also the effective area) for angles of incidence for which the metamaterial surface is located between the ground plane and the source. However, this comes at the cost of directionality since the ground plane will reflect away any signals coming from angles of incidence for which the ground plane is located between the rectenna and the source.

The point about the ground plane increasing the directivity is important, because it may not be feasible to build an omnidirectional energy harvesting device capable of collecting enough ambient RF power to operate practical devices. Energy harvesting efficiency is proportional to the effective area, which is a function of wavelength and directivity. Since the wavelength will be pre-determined by the source from which energy is being harvested, the only way to increase the effective area is by increasing the directivity. However, it is impossible to simultaneously increase the directivity in all directions because an increase in directivity for some directions necessitates a decrease in others. To demonstrate the difficulty of energy harvesting with a low-directivity antenna, consider a perfectly efficient short dipole antenna with a directivity of 1.5; a well-known omnidirectional antenna. The total amount of energy that can be harvested at 900 MHz with a relatively high incident power density of 0.1 $\mu W/cm^2$ is about 13 μW. This is well below the 100 μW threshold that might be needed for typical devices as suggested by Visser[19]. In circumstances where the location of the ambient radiation source is known, it would be beneficial to instead use a highly directive receiving antenna aimed at the source to maximize the effective area.

Future improvements in the design will be aimed at increasing the effective area, and improving the rectification efficiency of the antenna. The effective area may be improved by considering alternate structures and geometrical arrangements. Alternate structures could allow for polarization-independent energy harvesting. It's possible that the SRRs in the sample may be placed too close together such that the capturing areas of adjacent rings overlap and thus reduce the amount of power captured by each ring. A higher amount of power captured per ring should make it easier for the diodes to activate. Array theory for antennas suggests that the SRR's should be placed about half a wavelength apart for optimal power capture. To improve rectification efficiency, it may be worthwhile to replace the Schottky diodes with spin diodes[20] due to their superior zero-bias resistance. Another possibility is to replace the diodes with more complicated rectifying circuits that will increase DC voltage such as the Greinacher circuit[17].

Conclusion

Designing RF harvesting rectennas based off of metamaterial perfect absorbers is a promising solution for collecting ambient RF energy in low power density environments because they are tunable, highly efficient, and electrically small. However, to operate practical devices, the efficiency needs to be improved further, particularly in regards to overcoming the zero-bias resistance to activate the diodes. The reflecting ground plane component of the metamaterial rectenna dramatically improves the RF-DC conversion efficiency, especially for lower available power densities, by helping to overcome the zero-bias resistance and by increasing the effective area of the antenna for certain angles of incidence. The

biggest disadvantage of the ground plane is that it precludes omnidirectionality, although this may be unavoidable for effective energy harvesting.

Method

Fabrication

The metamaterial samples are fabricated using photolithography techniques. First, a mask that outlines a copper ring structure containing two gaps is designed and created. The dimensions of the unit cell and the design of the complete sample are shown in figure 1. The mask is placed on top of a copper-plated piece of printed circuit board composed of FR4 substrate which is then placed under an exposure lamp for 10 minutes. The mask is removed and the sample is washed with a diluted developer solution and then bathed in ferric chloride until any copper that was not shielded by the mask has been removed. After the excess copper has been removed, Skyworks SMS 7630 Schottky diodes are soldered across the right-hand gap of each ring with the orientations indicated by the red triangles on the unit cell shown in figure 1. The copper rings in the metamaterial capture RF waves, which are rectified by the diodes to deliver DC power to a load. The configuration of an antenna with a built-in rectification circuit is commonly referred to as a rectenna.

It is important to note that the rectenna sample is very sensitive to the polarization and angle of incidence of incoming RF signals. Since the sample is composed of an array of copper loops, electrical current can be driven by both the electric and magnetic fields. The orientation of the fields with respect to the sample determines which resonance modes will be excited and at which frequency they will occur. As such, deviations in the angle of incidence and/or polarization affect both the amount of energy harvested and the optimum frequency.

Experimental Setup

A signal generator produces a sinusoidal RF signal and sweeps both frequency and output power with frequencies ranging from 0.7-2.0 GHz. The signal is sent to an amplifier which increases the signal strength by about 45 dB (depending on the frequency). A small fraction of the amplified signal is extracted with a coupler, and then measured with a signal analyzer to determine the strength of the amplified signal. The amplified signal is transmitted by an SAS-570 horn antenna at normal incidence to the rectenna sample located about 380 cm away from the transmitting antenna. A metal ground plane is placed behind the sample with the purpose of creating a Fabry-Perot cavity to increase the amount of captured RF radiation. The experiment takes place within an anechoic chamber to isolate the rectenna sample from multi-path interference and outside signals. The rectenna sample is connected to a decade resistor box, which is used as a proxy for a device that DC power will be delivered to.

Theory

The DC power (P_{DC}) harvested from the sample can be determined simply by measuring the DC voltage (V) across the load with a multimeter and applying the well-known equation $P_{DC} = V^2/R$. A program written with LabVIEW is used to sweep over the appropriate frequency and power ranges while recording the voltage across the load to measure the efficiency (η) as described below. The efficiency is measured by taking the ratio of the rectenna's effective area to its geometric area. Effective area is

defined as the ratio of the power absorbed at the load divided by the Poynting vector at the location of the rectenna and is given by[21]:

$$\frac{P_{DC}}{S} \tag{1}$$

To determine the poynting vector at the location of the rectenna sample, the following equation from antenna theory is used[21]:

$$S = \frac{P_t \times G_t}{4\pi D^2} \tag{2}$$

P_t is the power radiated from the transmitting antenna, G_t is the linear gain of the transmitting antenna, and D is the distance between the transmitting antenna and the sample. The gain of the transmitting antenna was determined by using the standard 3-antenna technique for gain measurements[21].

Putting this all together leads to the following definition which will be called the energy harvesting efficiency:

$$\eta = \frac{\frac{V^2}{R}}{S \times A_{geo}} \times 100\% \tag{3}$$

This is actually the same way that aperture efficiency is defined for aperture antennas. For aperture antennas the efficiency will always be less than 100%, but for non-aperture antennas the effective area can be larger than the geometric area (a classic example of this is a dipole antenna). In such cases this definition of efficiency can yield results greater than 100%. It is important to recognize that this would not imply that the receiving antenna is capturing more power than is incident on it.

To characterize the efficiency of an energy harvesting system completely, the RF to DC conversion efficiency ($\eta_{RF\text{-}DC}$) should also be determined. $\eta_{RF\text{-}DC}$ indicates how much of the captured energy was successfully converted to DC power. This is the ratio of the DC power harvested to the RF power captured by the antenna.

$$\eta_{RF-DC} = \frac{P_{DC}}{P_{RF}} \times 100\% \tag{4}$$

Due to the nonlinearity of the diodes contained within the rectenna, a way to determine P_{RF} has not been determined, and so the focus of this paper is only on measuring the energy harvesting efficiency, η. Methods for determining P_{RF} will be explored in future experiments.

Experimental Concerns

There are a few considerations concerning the accuracy of the data. First, standing waves are generated between the transmitting antenna and the sample. This will result in higher power transfer at some frequencies and lower power transfer at others for a given distance between the sample and the transmitting horn antenna. It's estimated that this affects the power received by the sample varying by as much as ±10%. However, this effect is minimal at the frequency for which the power transfer is most efficient, since this is when the sample's reflection coefficient is at its smallest. Hence, the measurements taken at the frequency of peak efficiency will be the most accurate. Second, there is some power transmitted to the sample due to a second harmonic generated by the amplifier used for the transmitting antenna. The harmonic transmits a maximum of 5% of the total power radiated from the transmitting antenna, which occurs at the largest power densities that can be produced by the system. However, the harmonic signals are of frequencies that are inefficiently absorbed by the sample and thus are a very small contribution to the absorbed power. This effect is essentially negligible even at the highest measured power densities. Lastly, the precision of the transmitting system is not as good at high power due to limitations of the power settings of the signal generator. This results in some efficiency measurements appearing artificially lower than they should be. This effect is negligible when the power density is under 35 $\mu W/cm^2$. It is also negligible for power densities up to 50 $\mu W/cm^2$ while the frequency is less than about 1.3 GHz.

Some simplifications to this experiment have been made that slightly reduced the accuracy of the results, but greatly eased taking measurements. Firstly, since the impedance of the diodes is a function of both frequency and power, the impedance of the load should be adjusted accordingly for each measurement to maximize the power transferred to the load. We found that when the load was set to 1000 ohms, that the deviations from optimal efficiency were very small with changes in frequency and power. Secondly, the inclusion of a power management circuit might be necessary for real world applications to compensate for fluctuations in the signal being harvested. This is left out of the experiment so that we can focus on the energy harvesting efficiency of the rectenna alone.

Simulations

To supplement the experimental results, the absorption properties of the rectenna are simulated using CST Microwave Studio. Limitations in integrating circuit models into CST for plane wave simulations restricted us to using the Schottky diode model shown in figure 1, which is a simplified version of one used by Keyrouz et al[18]. The parameters for the model were obtained from the manufacturer's specifications. A broadband radar cross section monitor is used during a transient solver simulation of a plane wave to measure the absorption cross section (ACS) of the rectenna, which gives a rough indication of which frequencies the energy harvesting efficiency will be best. For the ACS calculation, the idealized diode in the equivalent circuit model is omitted to avoid generation of harmonics that interfere with the results. A surface current monitor is then used to identify the resonance modes of the rectenna at its peak energy harvesting frequency. Simulating the absorption cross section is primarily useful for estimating the peak frequencies for energy harvesting, but it is not good for predicting the correct proportionality for the efficiency across all frequencies. This is partly because of the limitations of the simple diode model used in the simulation. Another reason that the proportions may differ from measurements is that the absorption cross section includes energy losses in the rectenna, whereas the efficiency refers only to power absorbed in the load. As an additional test for energy harvesting capability of the sample, an incident Gaussian pulse was simulated and power delivered to a load was calculated. Although this technique will not indicate which frequency is best for producing DC power, it helps for getting an idea about how well the sample rectifies captured waves.

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innovating communications

Signal Optimization and Rectenna Design for Electromagnetic Energy Harvesting and Wireless Power Transfer

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Outline

- Introduction
- Rectenna design
  - Dual band
  - Load independent performance
- Signal design
  - Random modulation, noise, chaotic signals
  - Mode locked oscillators
- Some circuits…!
  - Solar powered RF circuits
  - Multi-technology harvesters
- Conclusion

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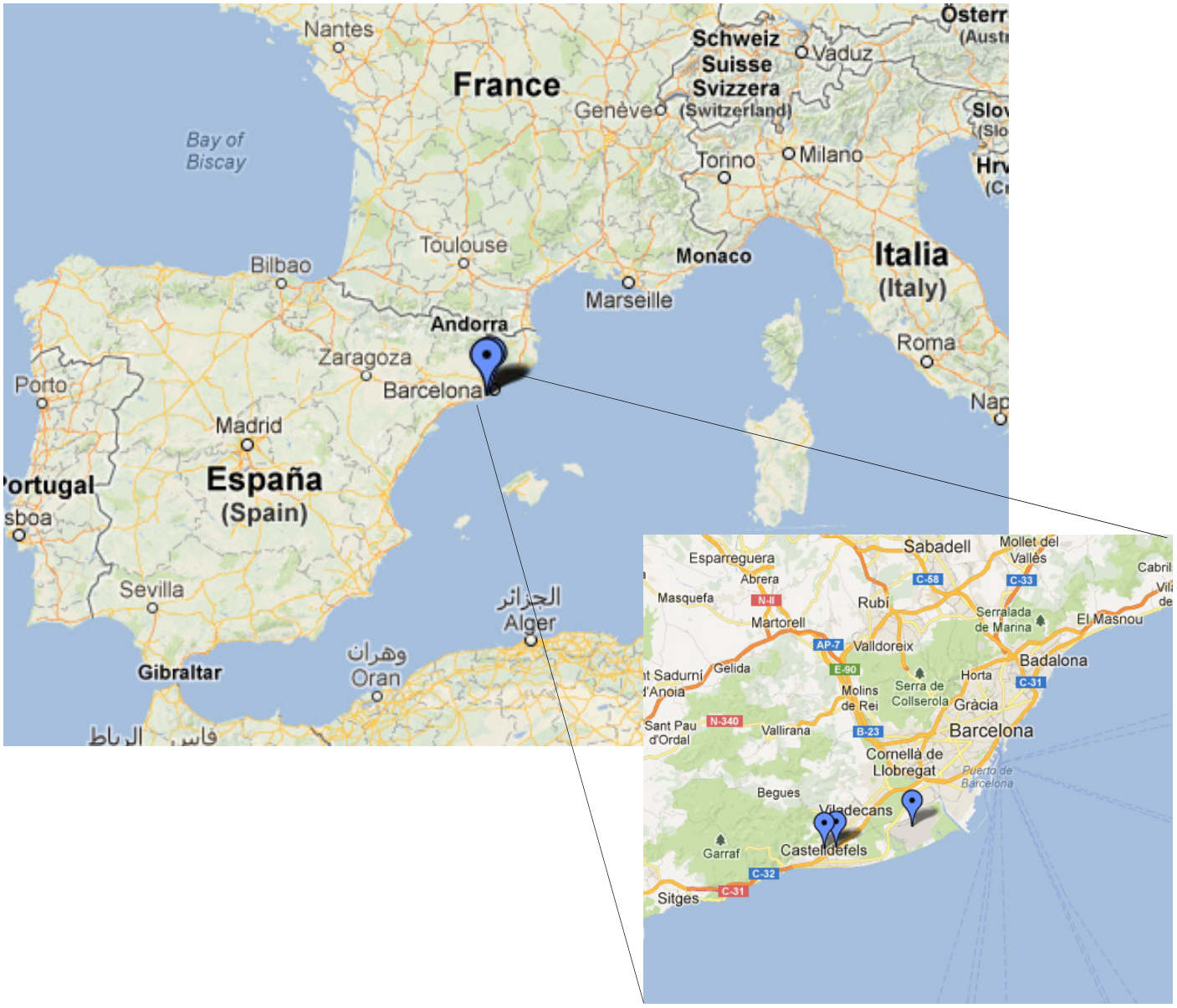
CTTC, Castelldefels – Barcelona

Founded in 2001



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CTTC, Castelldefels – Barcelona

- Research staff: 35 Ph.D., 20 M.Sc, 3500-m^2 building
- 3 Research Divisions: Comm. Systems, Comm. Networks, Comm. Technologies
- ***Department of Microwave Systems and Nanotechnology***

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CTTC, Castelldefels – Barcelona, SPAIN

Active microwave circuit design

- **Energy Harvesting and RFID**
- Oscillator design including integrated CMOS oscillators (Fig. 1)
- Active antennas, phased arrays (Fig. 2), retro-directive arrays (Fig. 3)
- Substrate Integrated Waveguide (SIW) (Fig. 4)
- Efficient Power Amplifier (Fig. 5)

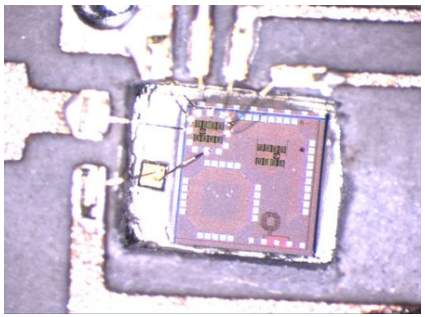


Fig. 1. CMOS VCO for UWB-FM

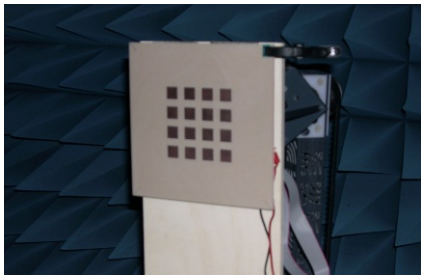


Fig. 2. C-band Coupled Oscillator Reflectrarray prototype

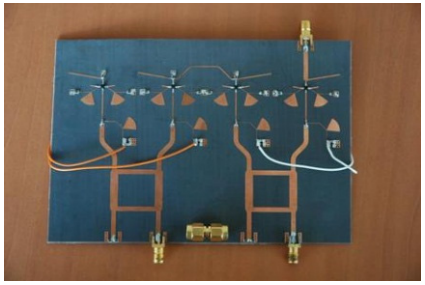
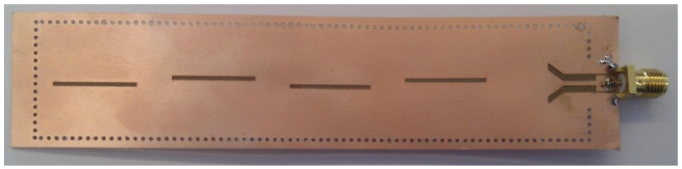
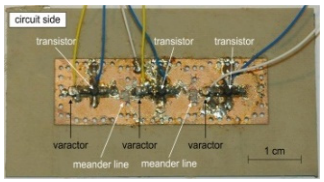


Fig. 3. S-band retro-directive array.





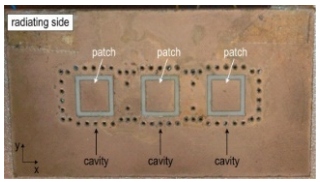


Fig. 4. SIW circuits.

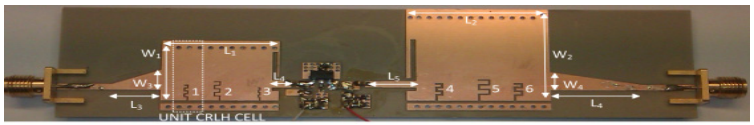


Fig. 5. Power Amplifier (SIW).

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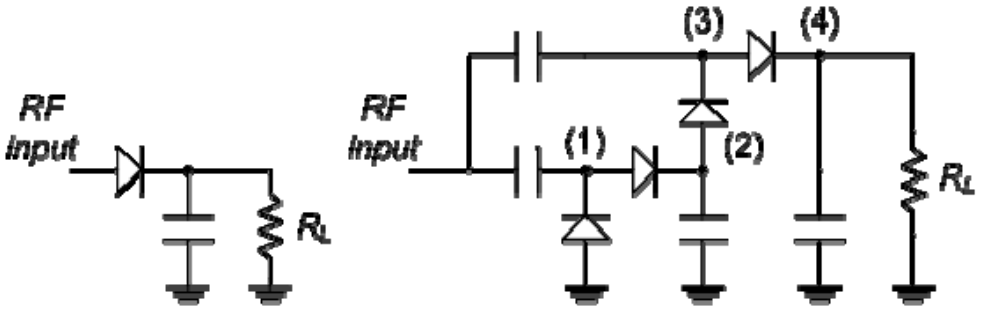
Rectenna Design

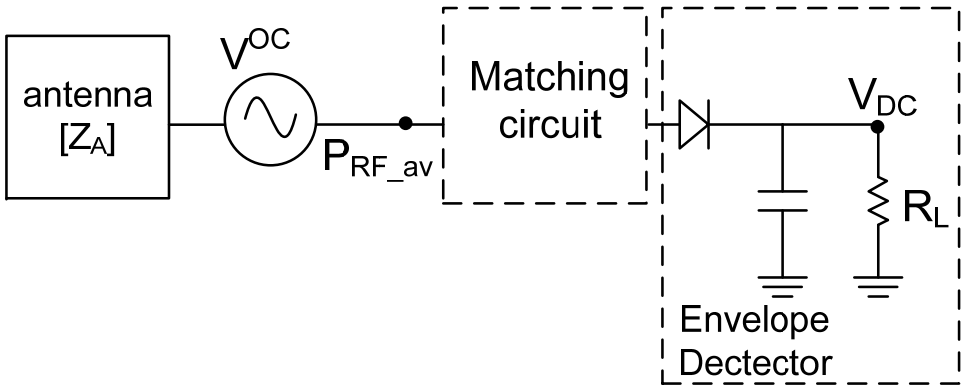
- Rectifier circuits: envelope detector, charge pump circuits
- Schottky diodes, low / zero barrier diodes



Reported UHF rectifier efficiencies for available input power levels in the order of 10 uW are near 20%, and increase to >50% for available power levels of 100uW.

Rectenna Design

- Rectenna optimization using the RECEIVE antenna Thevenin (or Norton) equivalent circuit



- Multiple goal harmonic balance for optimizing the RF-DC conversion efficiency

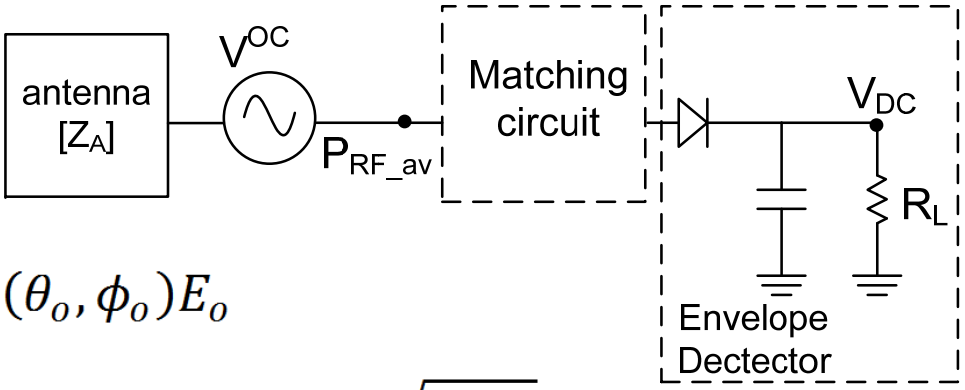
$$\eta = \frac{P_{DC}}{P_{RF,av}} = \frac{V_{DC}^2}{P_{RF,av} R_L}$$

Georgiadis, A.; Andia Vera, G.; Collado, A., "Rectenna design and optimization using reciprocity theory and harmonic balance analysis for electromagnetic (EM) energy harvesting," *Antennas and Wireless Propagation Letters, IEEE* , vol.9, no., pp.444,446, 2010

Rectenna Design

- Open circuit voltage maybe calculated using reciprocity theory

$$\eta = \frac{P_{DC}}{P_{RF,av}} = \frac{V_{DC}^2}{P_{RF,av} R_L}$$



$$V_{V,H}^{oc}(\theta_o, \phi_o, S) = \frac{4\pi}{jk\eta_o} F_{V,H}(\theta_o, \phi_o) E_o$$

$$E_o = \sqrt{2\eta_o S}$$

$$P_{av} = \frac{1}{4} V^{oc\,*} [Z_A + Z_A^*]^{-1} V^{oc}$$

$$\eta_o = 120\pi$$

- Harmonic balance for the optimization of the RF-DC conversion efficiency

Georgiadis, A.; Andia Vera, G.; Collado, A., "Rectenna design and optimization using reciprocity theory and harmonic balance analysis for electromagnetic (EM) energy harvesting," *Antennas and Wireless Propagation Letters, IEEE* , vol.9, no., pp.444,446, 2010

Rectenna Design

- The open circuit voltage *MAGNITUDE* can also be calculated using the antenna effective area

$$E_o = \sqrt{2\eta_o S}$$

$$V_{V,H}^{oc}(\theta_o, \phi_o, S) = \frac{4\pi}{jk\eta_o} F_{V,H}(\theta_o, \phi_o) E_o$$

$$A_e = \frac{\lambda^2}{4\pi} G \qquad P_{av} = A_e S$$

$$P_{av} = A_e S = \frac{\lambda^2}{8\pi\eta_o} G(\theta_o, \phi_o) E_o^2$$

HOW DO THEY COMPARE ?

$$V_{oc} = \sqrt{8 P_{av} R_A} = \sqrt{\frac{G(\theta_o, \phi_o) R_A}{\pi\eta_o}} \lambda E_o$$

C. A. Balanis, Antenna Theory: Analysis and Design, 3rd Ed., Wiley,2005
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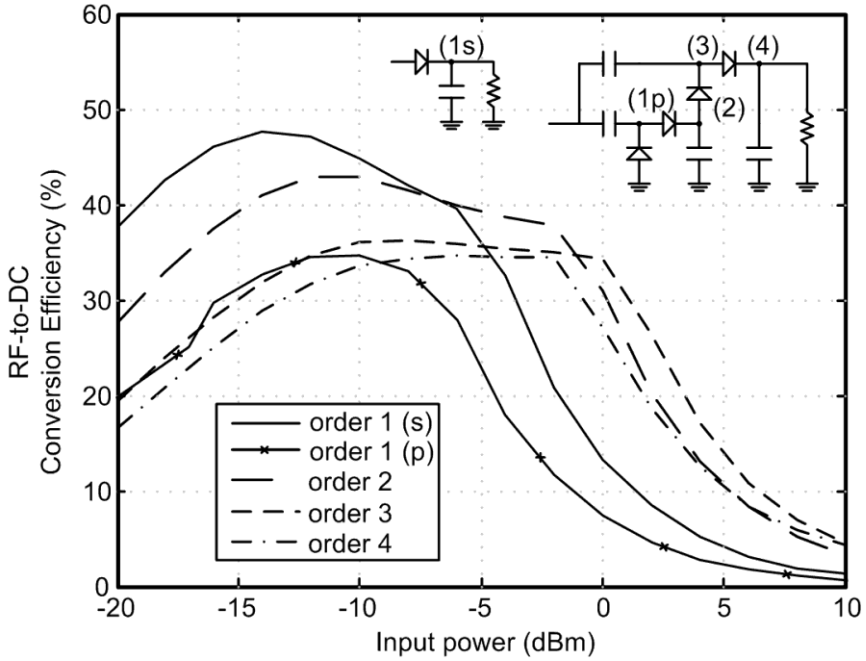
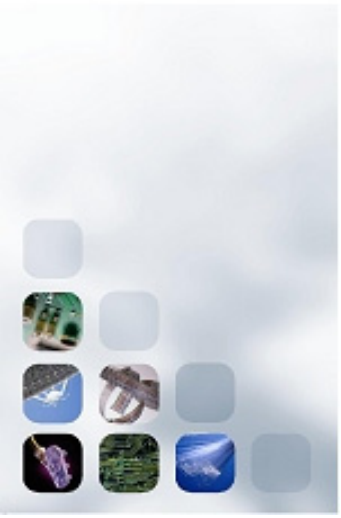
Rectenna Design

- Circuit topology important in low available power conditions
- Trade-off between efficiency and output voltage





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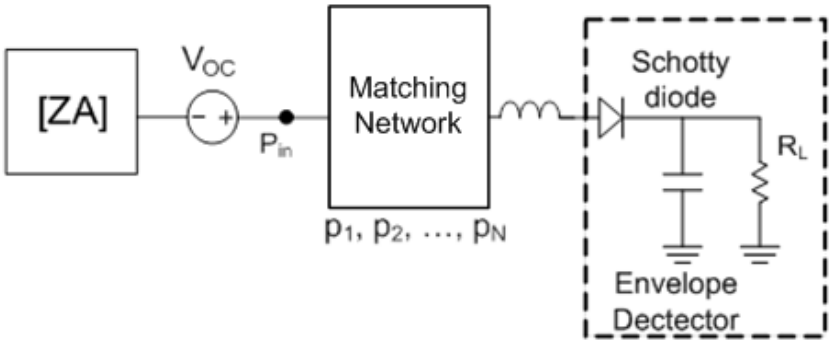
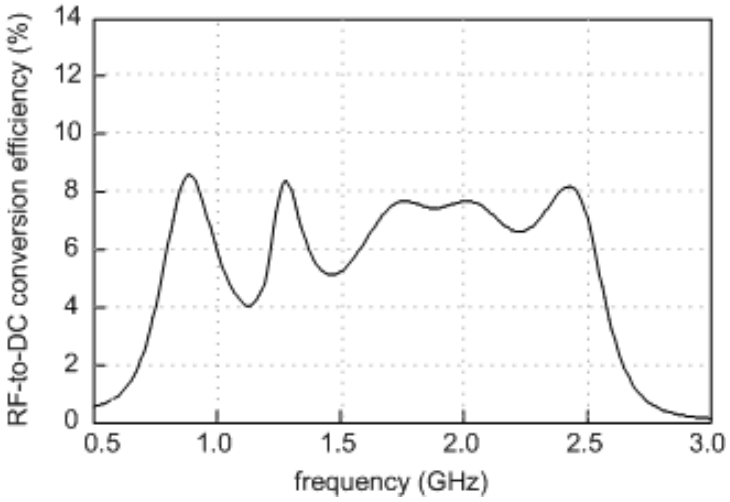
RF to DC conversion efficiency optimization: Broadband Case





OPTIMIZATION PARAMETERS:

$p_1,\ldots,p_N$
R_L

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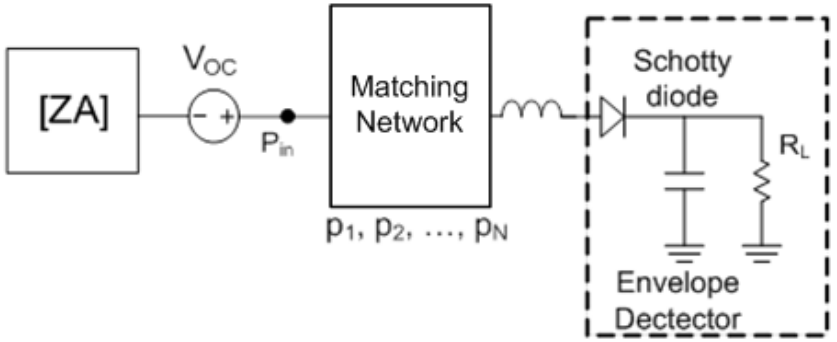
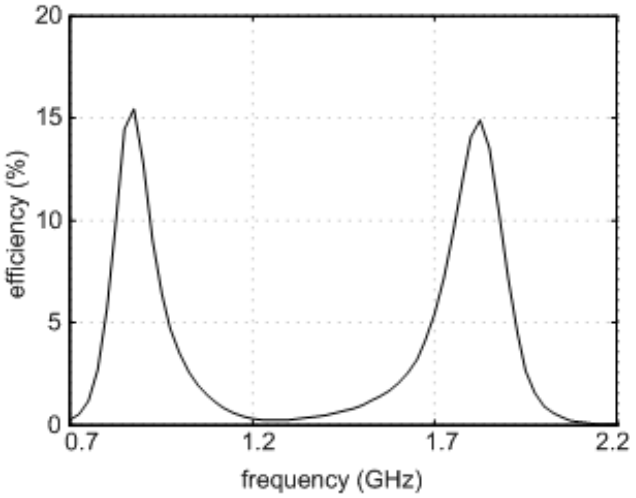
RF to DC conversion efficiency optimization: Dual-Band Case





OPTIMIZATION PARAMETERS:

$p_1,\ldots,p_N$
R_L

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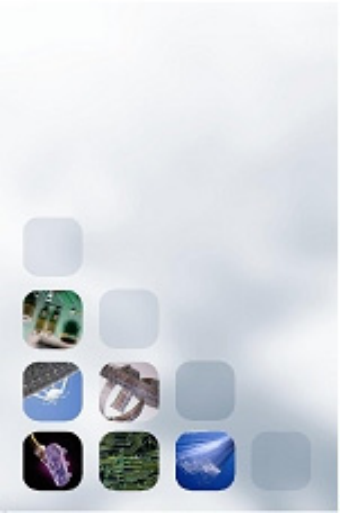
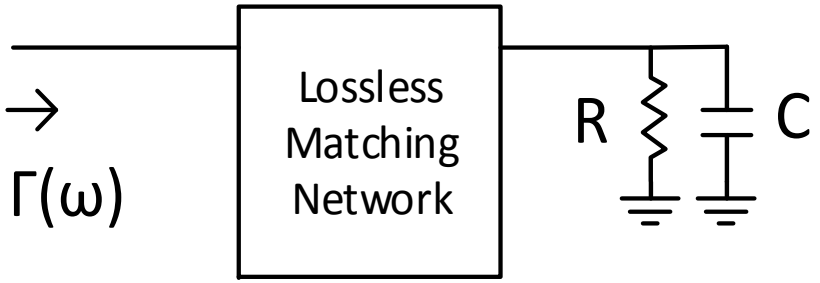


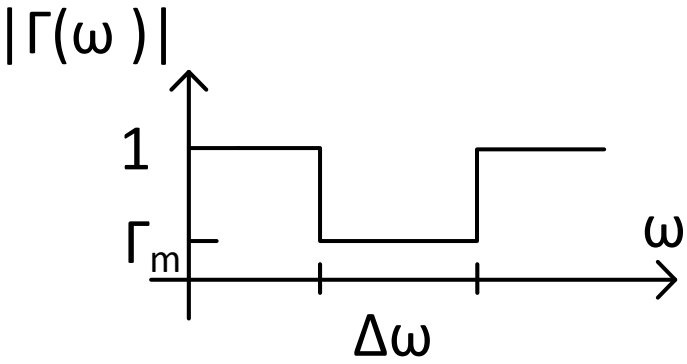
Theoretical limits: impedance bandwidth

Bode-Fano criteria *(see e.g. D. Pozar, Microwave Eng.)*

$$\int_0^{\infty} \ln \frac{1}{|\Gamma(\omega)|} d\omega \leq \frac{\pi}{RC}$$



$$\Delta\omega \ln \frac{1}{\Gamma_m} \leq \frac{\pi}{RC}$$



Let R = 1.5 KOhm, and C = 0.9 pF

What is the miminum reflection coefficient if one wants to cover a freq. band from 800 MHz to 2.6 GHz ?

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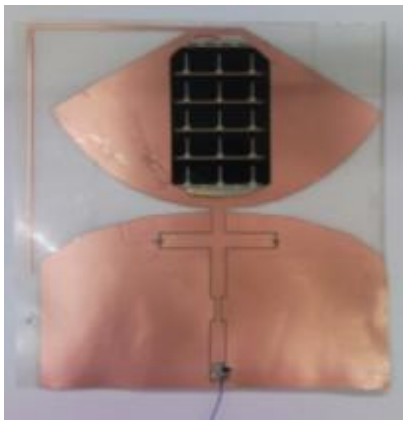
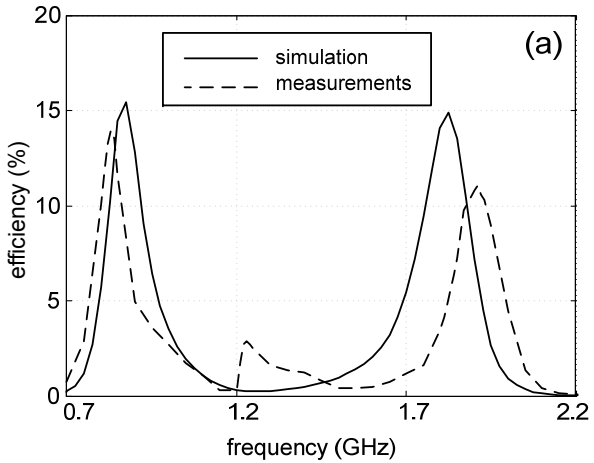
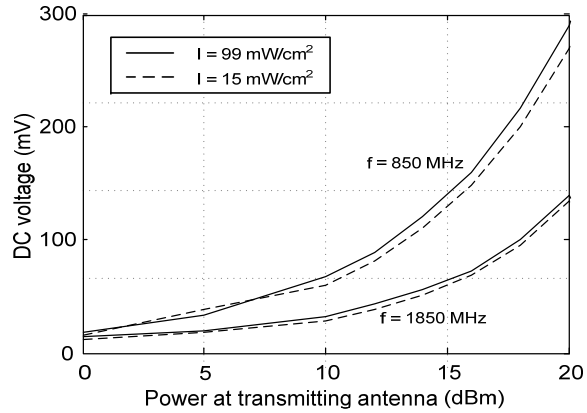
Rectenna Design

850 MHz/1850 MHz Dual Band Rectenna

- Broadband monopole antenna (0.7GHz - 6 GHz)
- Akaflex PCL3-35/75 μm with ε_r = 3.3 and tanδ = 0.08
- Silicon Schottky diode (Skyworks SMS7630)
- Coplanar waveguide matching network
- Optimization for input power of -20 dBm and R_L=2.2 kΩ







Collado, A.; Georgiadis, A., "Conformal Hybrid Solar and Electromagnetic (EM) Energy Harvesting Rectenna," *Circuits and Systems I: Regular Papers, IEEE Transactions on* , vol.60, no.8, pp.2225,2234, Aug. 2013

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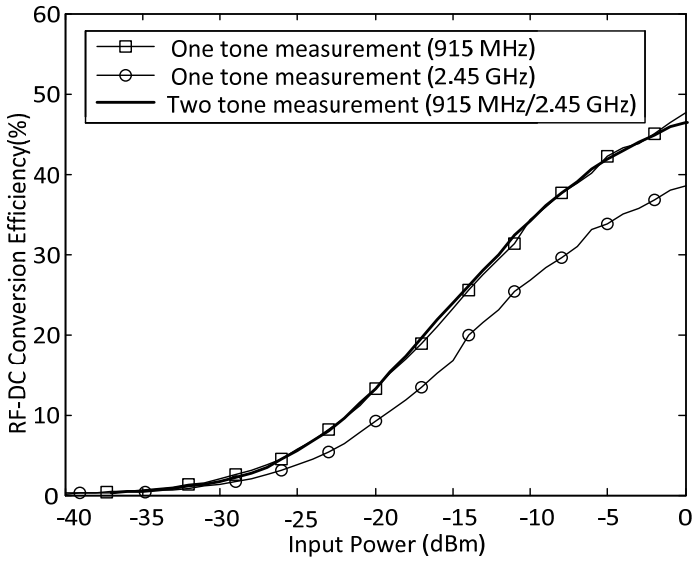
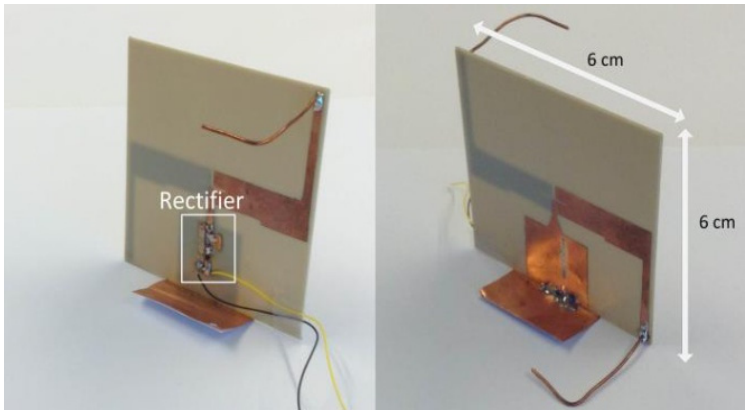
Rectenna Design

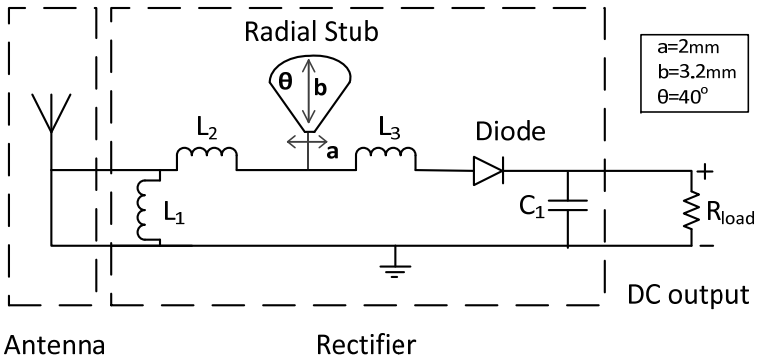
- Optimization goals are used to maximize the RF-DC conversion efficiency at 915 MHz and 2.45 GHz

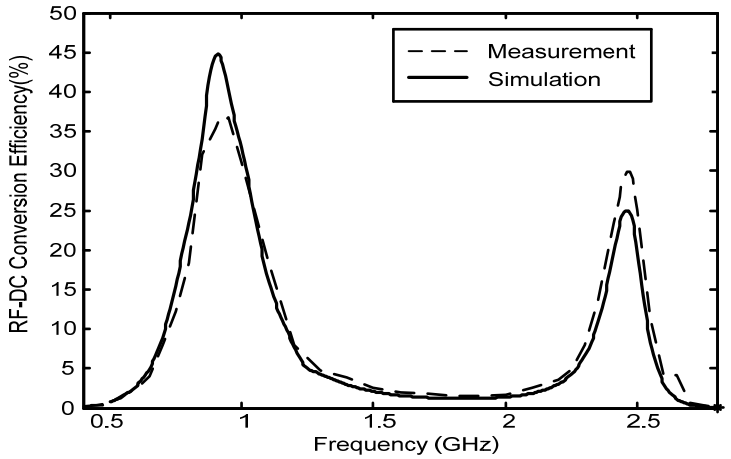
η = 48% and η = 39% at 915 MHz and 2.45 GHz, for P_{in}=0 dBm

η <1 % for P_{in}<-33 dBm









Niotaki, K.; Sangkil Kim; Seongheon Jeong; Collado, A.; Georgiadis, A.; Tentzeris, M.M., "A Compact Dual-Band Rectenna Using Slot-Loaded Dual Band Folded Dipole Antenna," *Antennas and Wireless Propagation Letters, IEEE* , vol.12, no., pp.1634,1637, 2013

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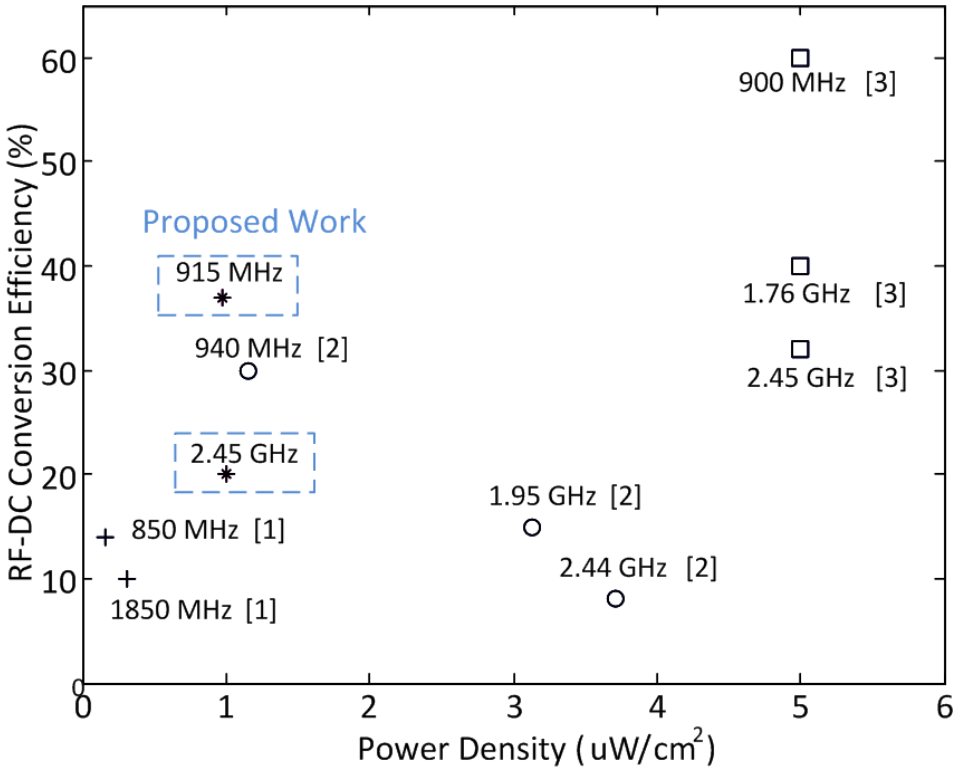
Rectenna Design



[1] A. Collado, and A. Georgiadis, "Conformal Hybrid Solar and Electromagnetic (EM) Energy Harvesting Rectenna," *IEEE Trans. Circuits Syst. I, Reg. Papers*, vol. 60, no. 8, pp.2225,2234, Aug. 2013

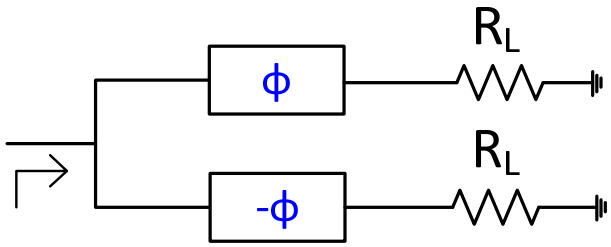
[2] B. L. Pham and A.-V. Pham, "Triple Bands Antenna and High Efficiency Rectifier Design for RF Energy Harvesting at 900, 1900 and 2400 MHz," in *Proc. IEEE MTT-S Int. Microwave Symp.*, Seattle, WA, 2–7 June 2013.

[3] V.Rizzoli, G. Bichicchi, A. Costanzo, F. Donzelli, and D. Masotti, "CAD of multi-resonator rectenna for micro-power generation," in *Proc. Microwave Integrated Circuits Conference (EuMIC 2009)*, 28-29 Sept. 2009, pp.331–334.

- η = 37% and η = 20% at 915 MHz and 2.45 GHz for a power density of 1 uW/cm^2
  - 1 uW/cm^2 corresponds to P_{in}=-9 dBm and P_{in}=-15 dBm at 915 MHz and at 2.45 GHz

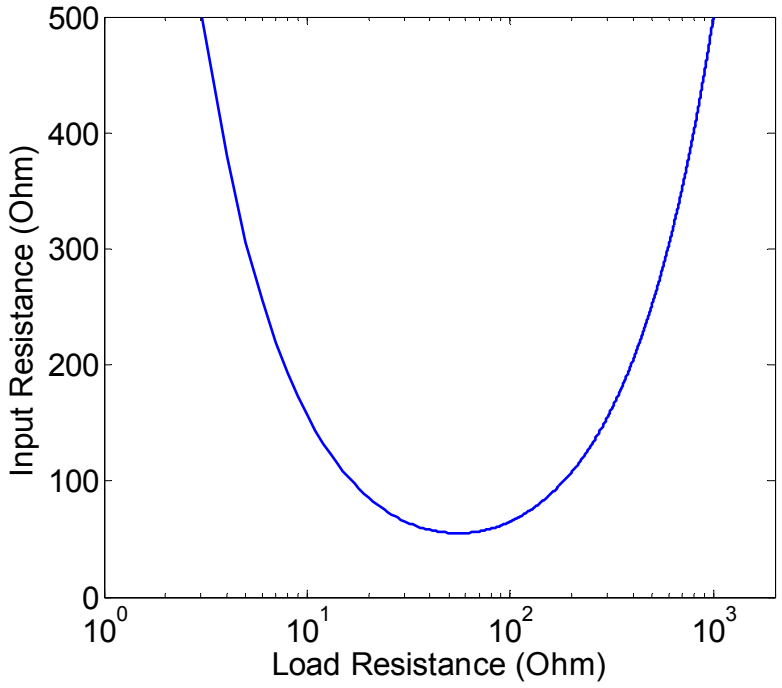
Rectenna Design

- Challenge: load and input power variation
- Resistance compression networks

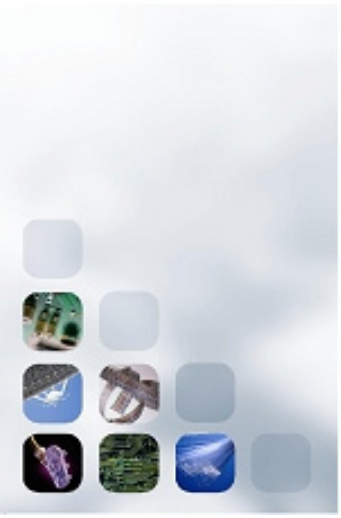


Load resistance variation: 3 Ohm – 1000 Ohm

Input resistance variation: 55 Ohm – 500 Ohm



Y. Han, O. Leitermann, D.A. Jackson, J.M. Rivas, and D.J. Perreault, "Resistance Compression Networks for Radio-Frequency Power Conversion, " *IEEE Trans. on Power Electronics*, vol. 22, no. 1, pp. 41-53, Jan. 2007.



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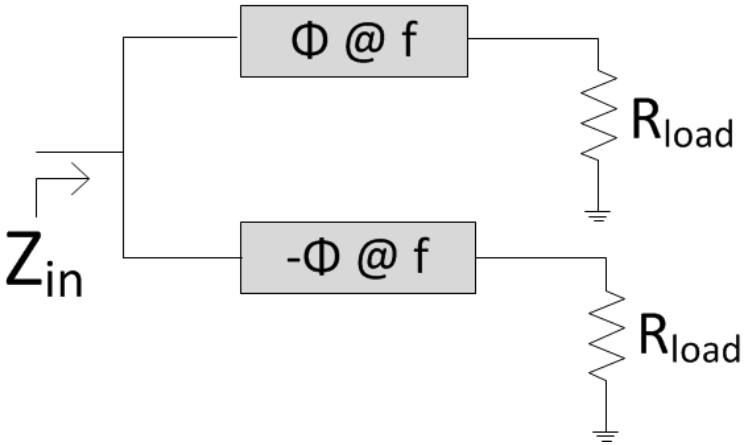
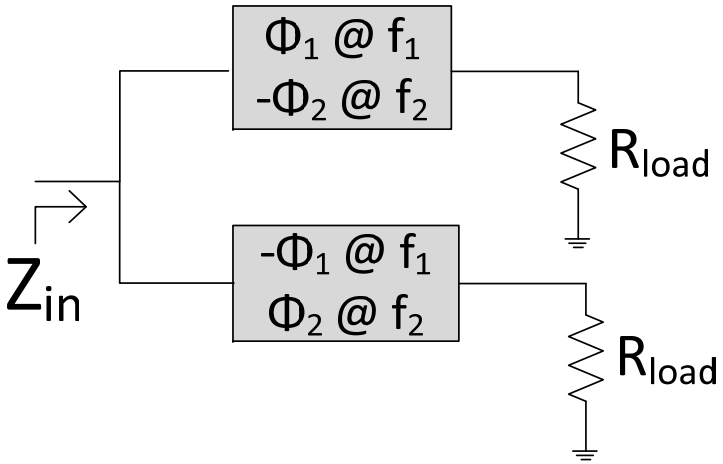
Dual-Band Resistance Compression Networks

- Resistance Compression Networks
  - Identical R_{load} variations
  - Opposite phase response



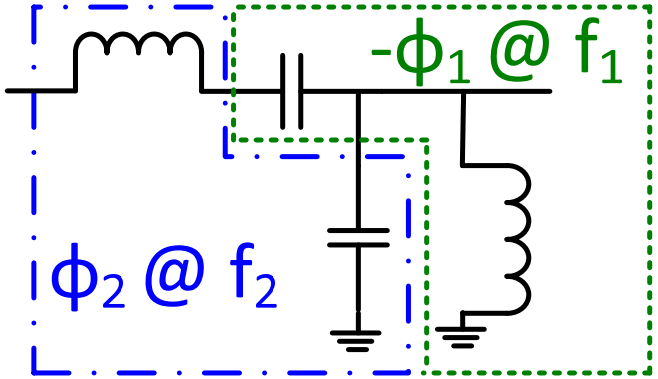
RCN operating at single frequency

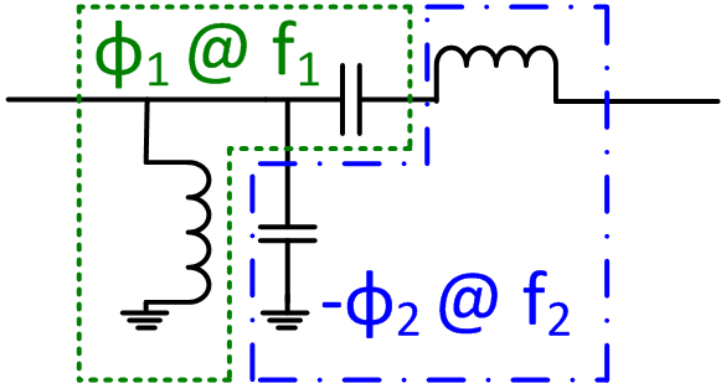


Dual-Band RCN

Dual-Band Resistance Compression Networks

- Dual-Band RCNs
  - **Opposite** phase response at f_1 and f_2 ($f_1 < f_2$)





→ $-\varphi_1$ at f_1, $+\varphi_2$ at f_2

→ $+\varphi_1$ at f_1, $-\varphi_2$ at f_2

K. Niotaki, A. Georgiadis, A. Collado, 'Dual-Band Rectifier Based on Resistance Compression Networks,' in Proc. IEEE MTT-S IMS, Tampa, 1-6 June 2014

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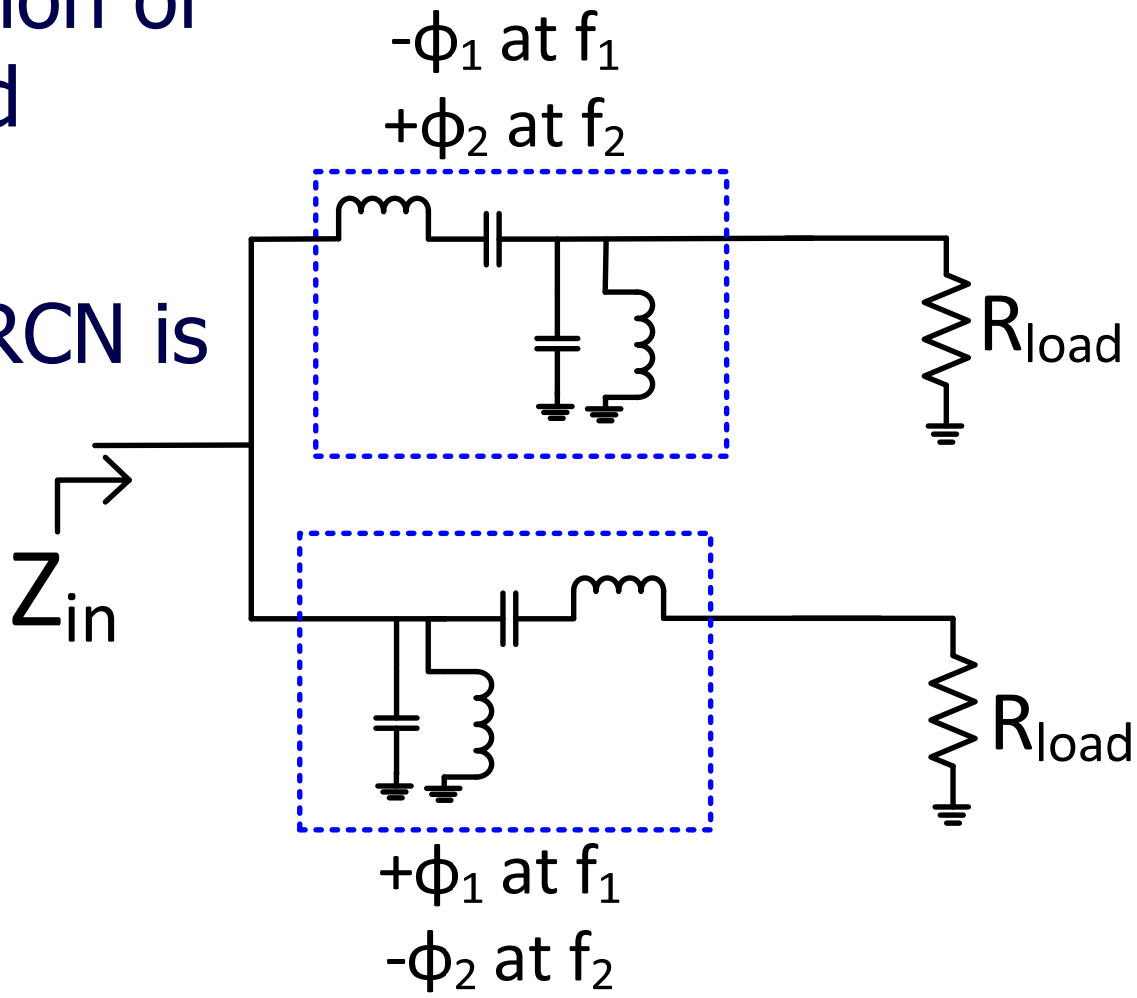
Dual-Band Resistance Compression Networks

- A small variation of Z_{in} is achieved
- A dual-band RCN is obtained



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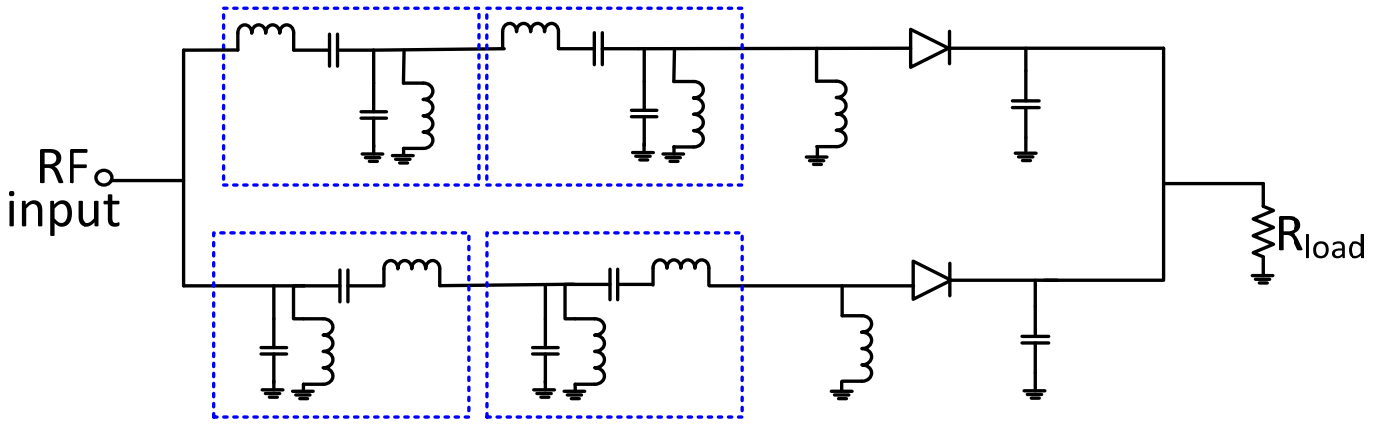
Dual-Band Rectifiers

- Dual-Band Resistance Compressed Rectifier
  - 2 unit cells for each branch
  - 915 MHz and 2.45 GHz
  - Skyworks Schottky SMS7630 diode



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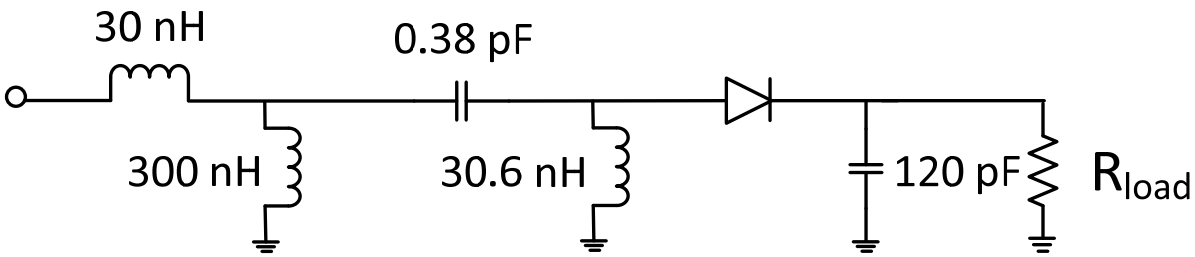
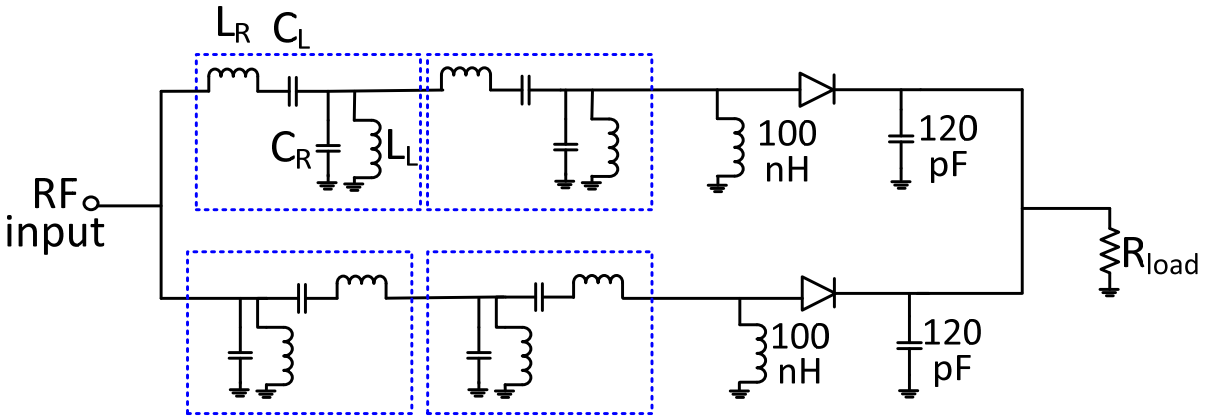
Dual-Band Rectifiers

- Performance comparison
  - Resistance compressed rectifier
  - Conventional envelope detector rectifier

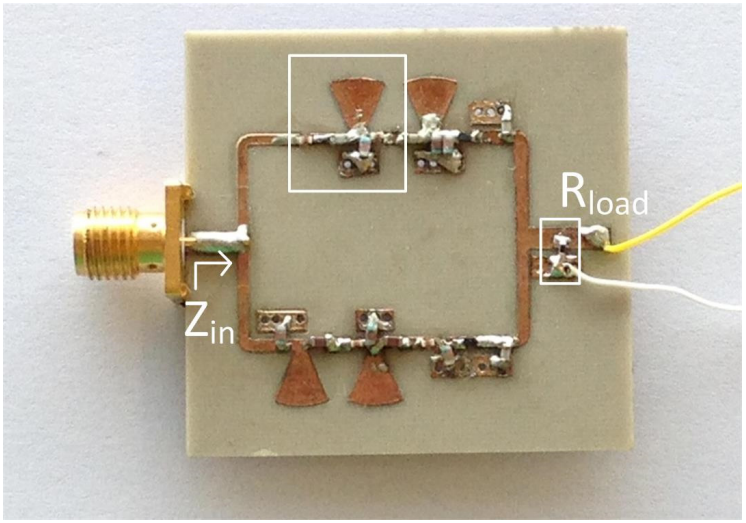


- Rectifier design
  - Harmonic balance analysis (*HB*)
  - Large signal Scattering parameters (*LSSP*)

Resistance Compressed Rectifier



L_R	8.7 nH
L_L	100 nH
C_R	0.8 pF
C_L	2.7 pF



- Schottky diode
  - SMS7630
- Arlon 25N
  - 30 mil
  - ε_r=3.38

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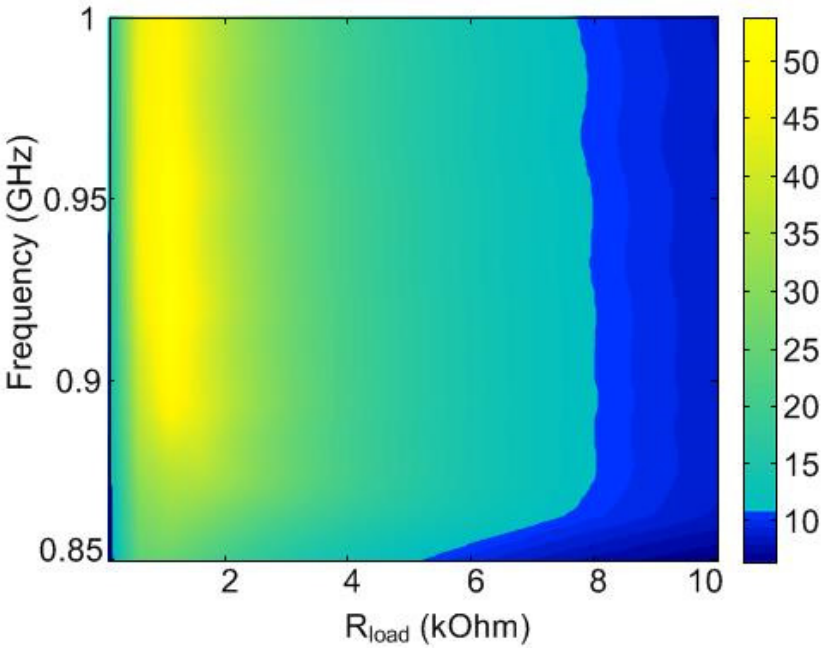
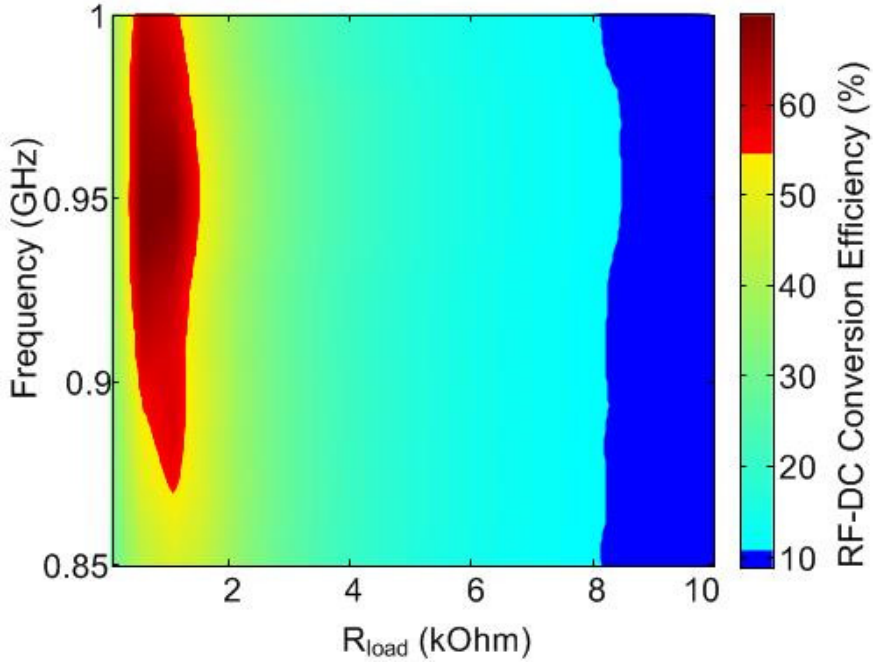
Dual-Band Rectifiers

- Efficiency Improvement of n
  - 37 % → 62.3 % for R_{load}=0.5 kOhm & Pin=0 dBm at 915 MHz



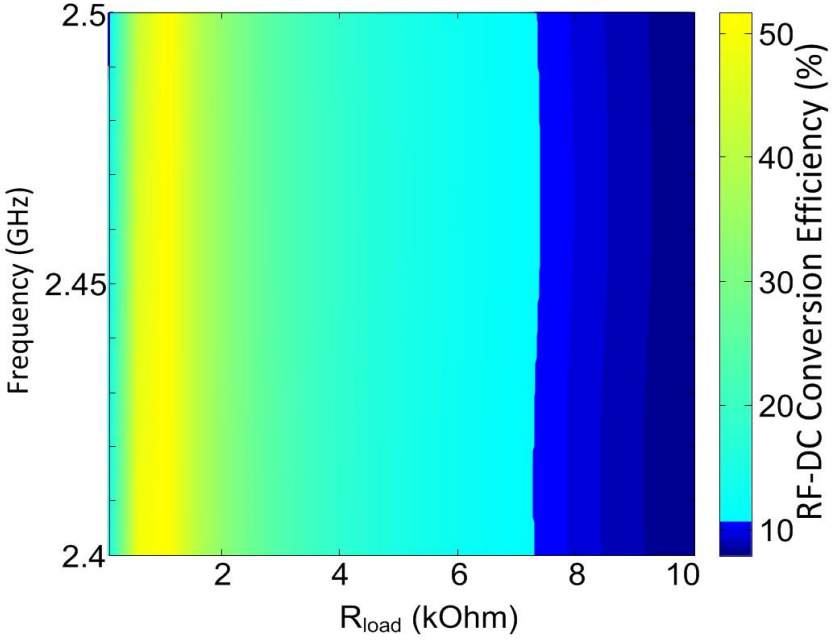
Conventional Rectifier



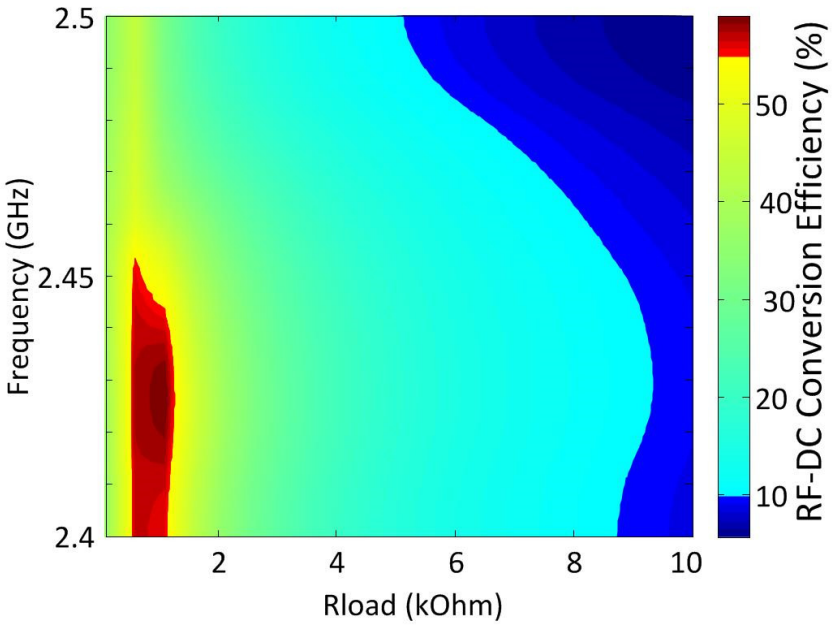
Resistance-Compressed Rectifier

Dual-Band Rectifiers

- Efficiency Improvement of n
  - 41.2 % → 54.4 % for R_{load}=0.5 kOhm & P_{in}=0 dBm at 2.45 GHz



Conventional Rectifier



Resistance-Compressed Rectifier

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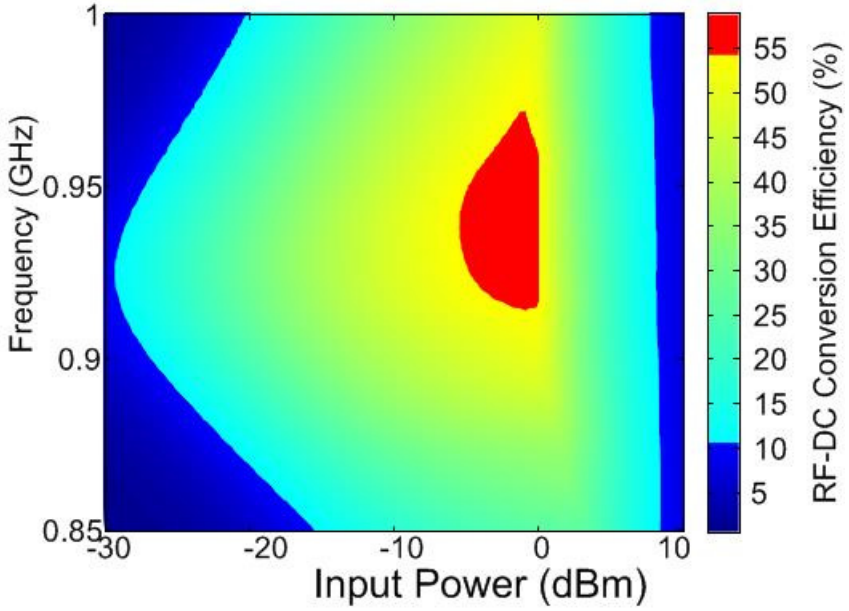
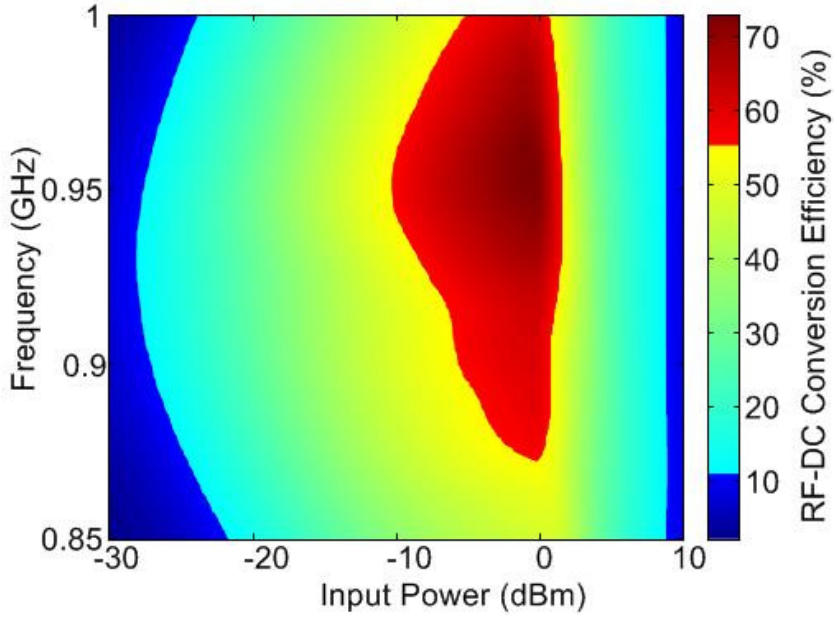
Dual-Band Rectifiers

- RF-DC conversion efficiency at 915 MHz
  - R_{load}=1 kOhm



Conventional Rectifier



Resistance-Compressed Rectifier

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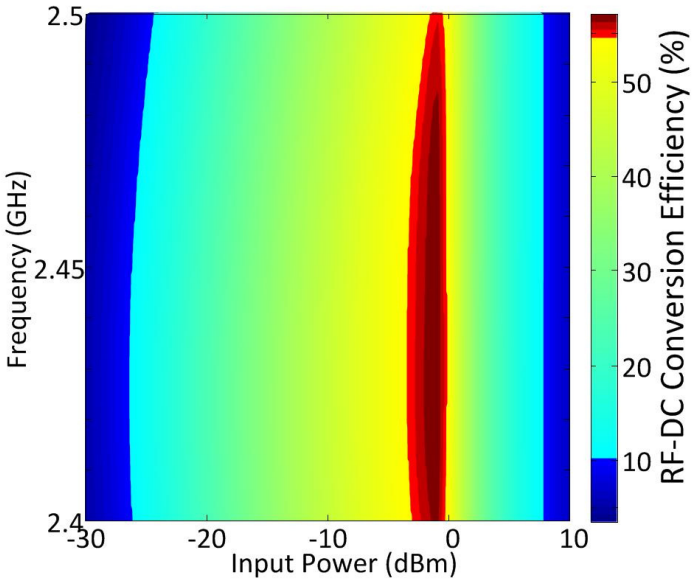
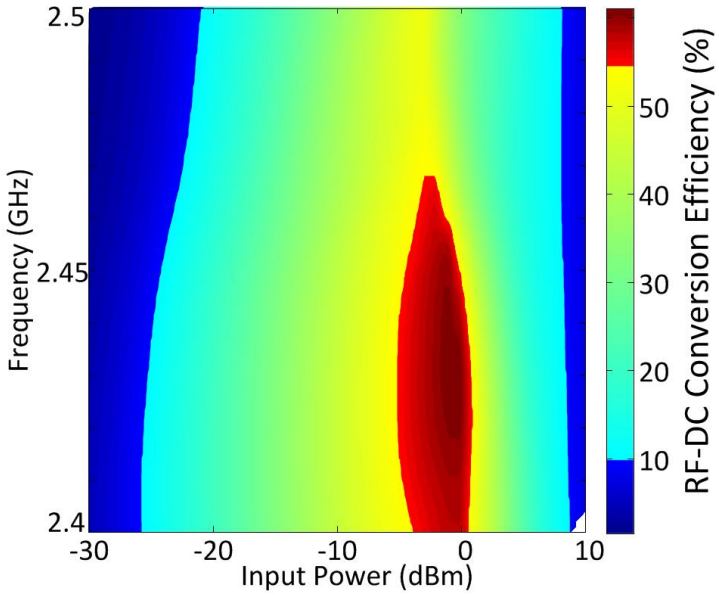
Dual-Band Rectifiers

- RF-DC conversion efficiency at 2.45 GHz
  - R_{load}=1 kOhm



Conventional Rectifier



Resistance-Compressed Rectifier

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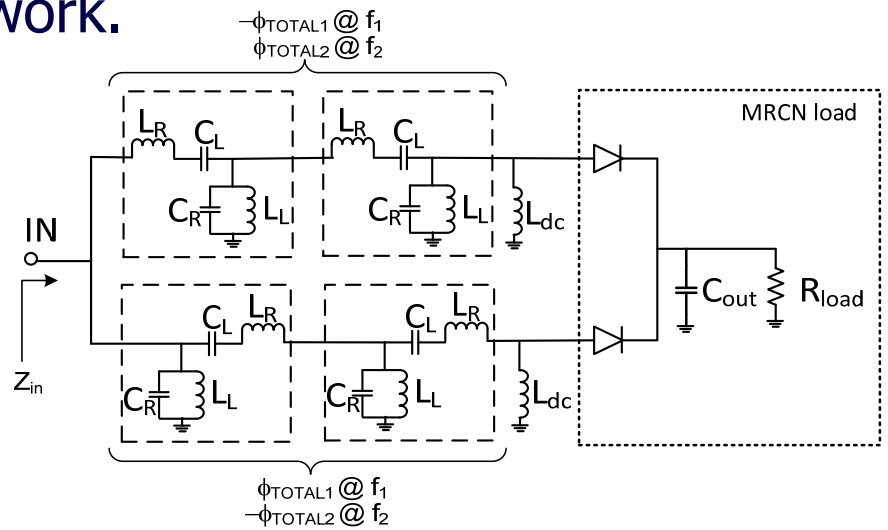
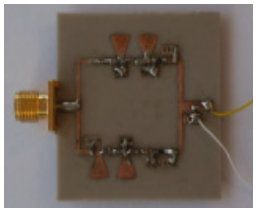
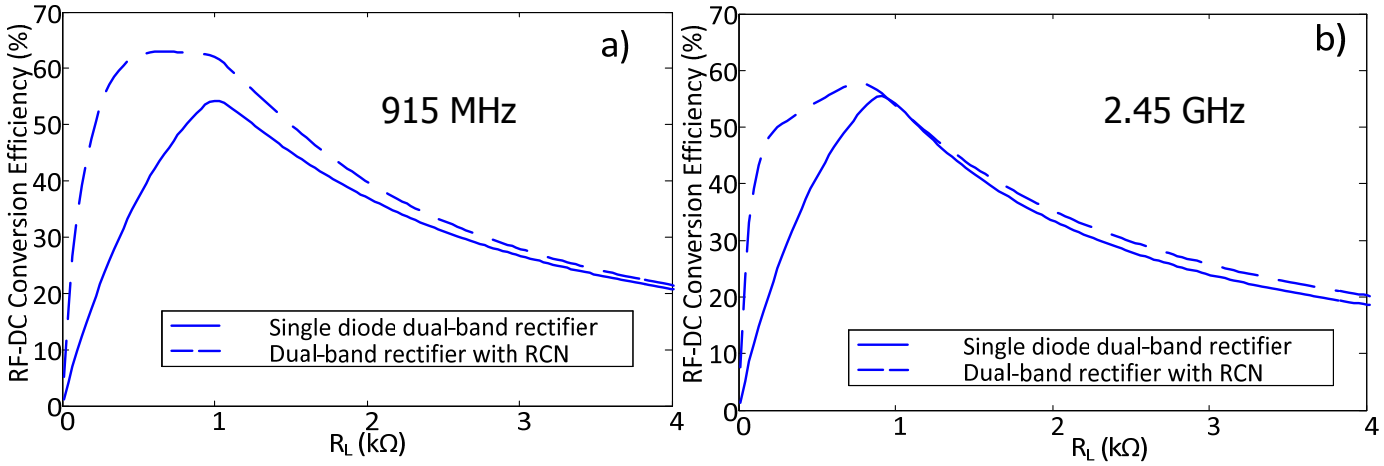
Rectenna Design

- Dual band metamaterial based resistance compression network.







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Signal Design

- Signals with time-varying envelope (PAPR > 0 dB) lead to higher rectifier RF-DC conversion efficiency
  - Multi-sines (Durgin, Carvalho, Popovic, …)
  - Chaotic signals
  - White noise
  - Random modulation (multi-carrier)

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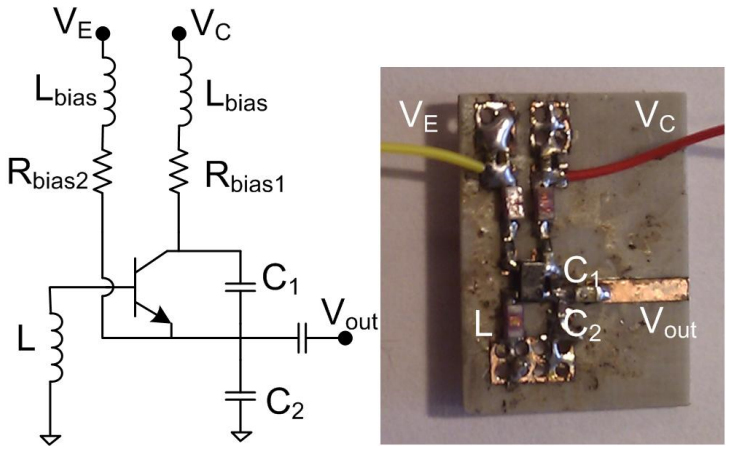
Signal Design

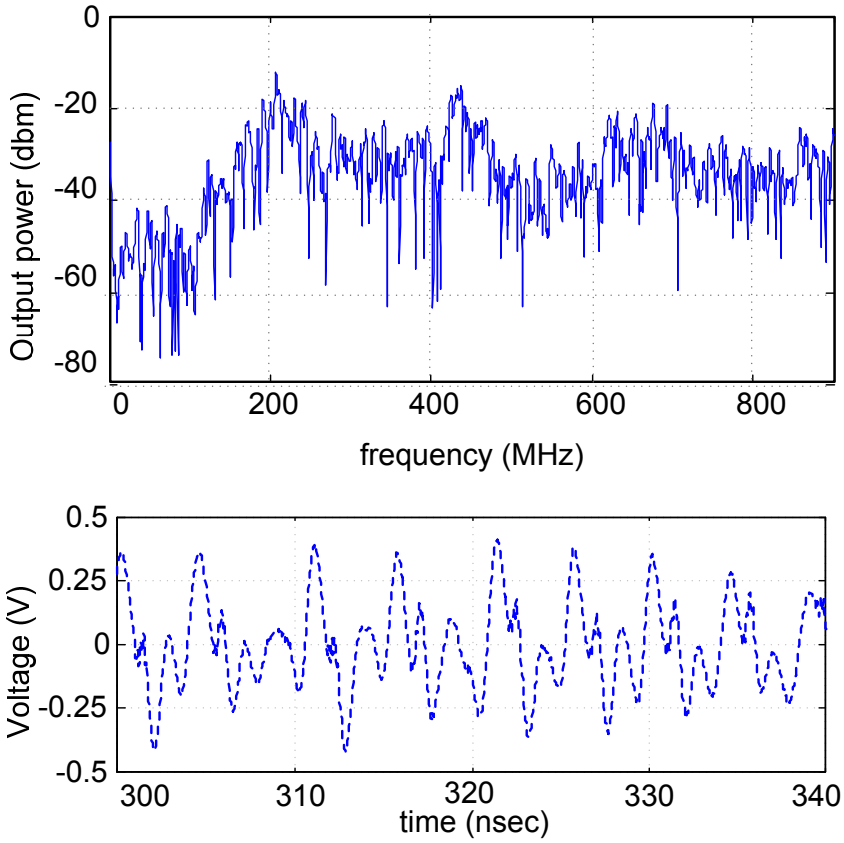
- First experiments: chaotic oscillator

❑ Colpitts based chaotic generator

❑ Bipolar transistor BFP183w



433 MHz chaotic generator



A. Collado, A. Georgiadis, "Improving Wireless Power Transmission Efficiency Using Chaotic Waveforms," in Proc. IEEE MTT-S IMS 2012, Montreal, 17-22 June 2012.

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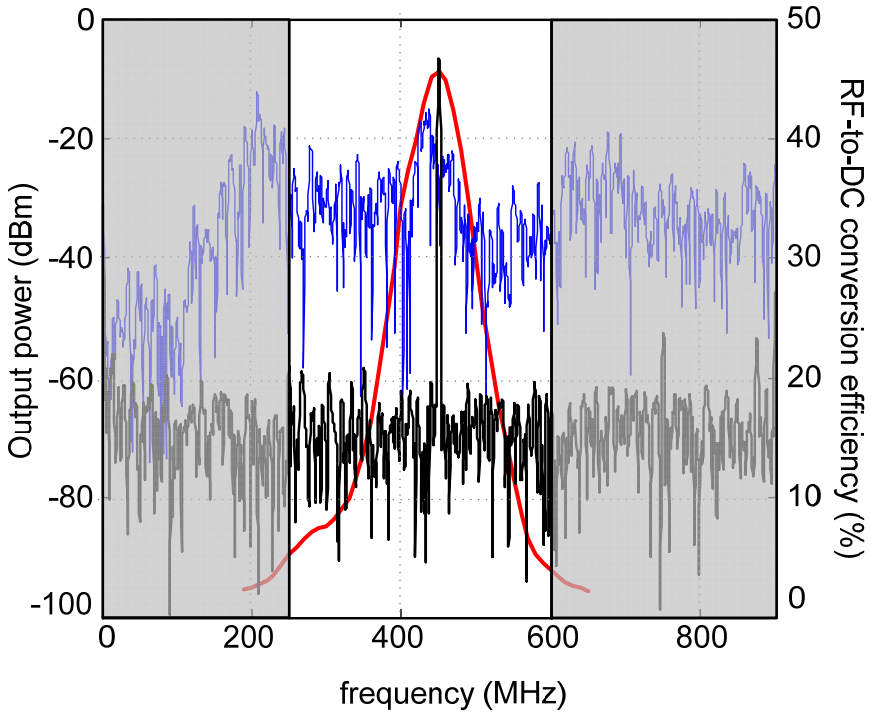
Signal Design

- Need to filter chaotic signal

chaotic signal power
[250 MHz – 600MHZ]

-6.5 dBm

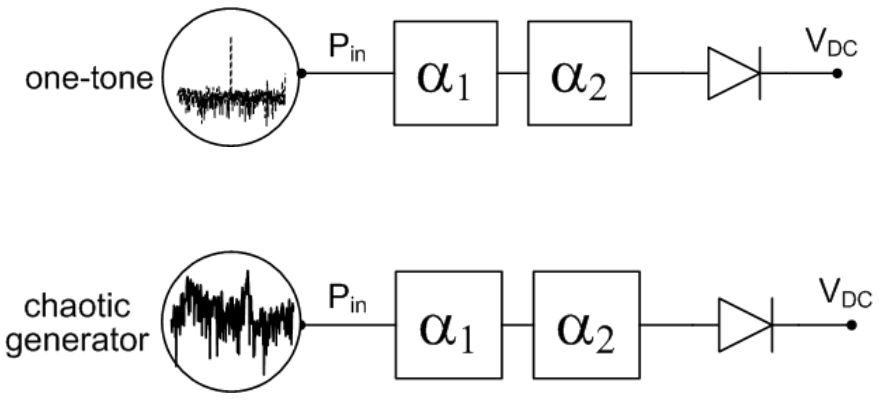
One-tone signal power
[250MHz – 600MHZ]

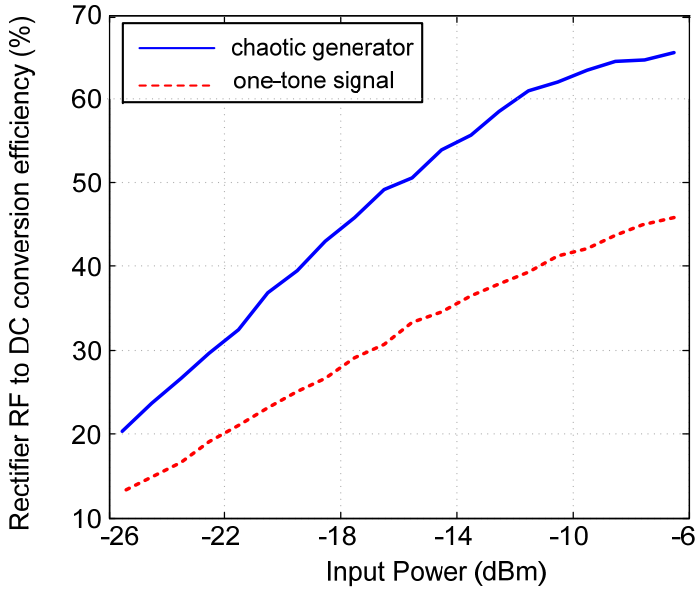


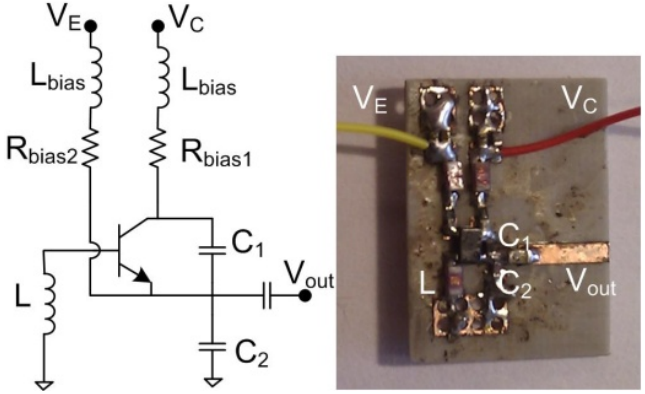
- Total power of 1-tone signal selected to be equal to the chaotic signal total power in the bandwidth of the rectifier

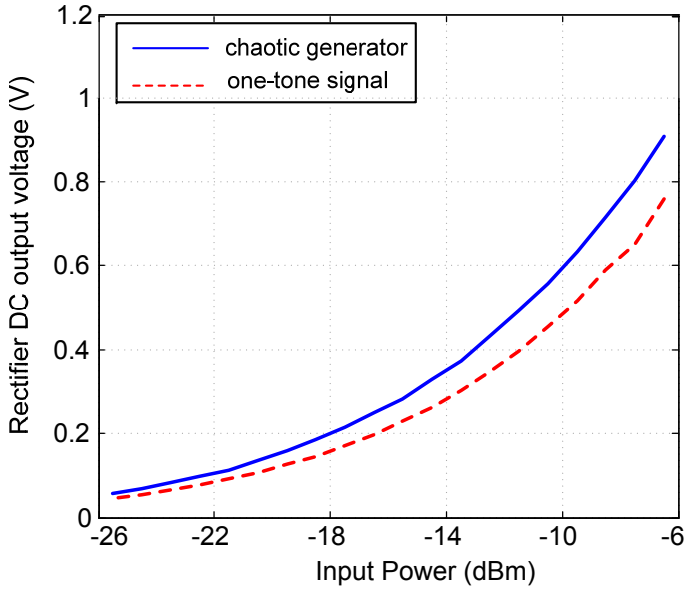
A. Collado, A. Georgiadis, "Improving Wireless Power Transmission Efficiency Using Chaotic Waveforms," in Proc. IEEE MTT-S IMS 2012, Montreal, 17-22 June 2012.

Signal Design









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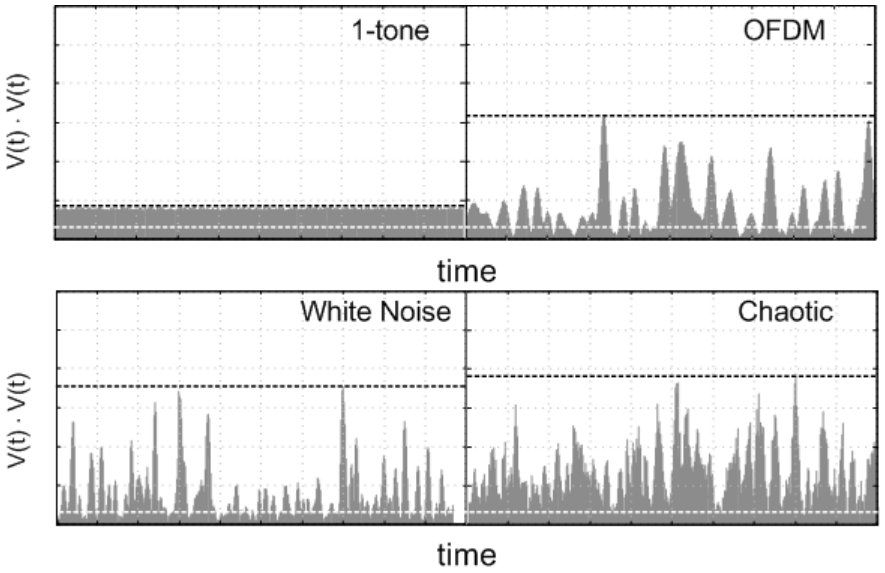
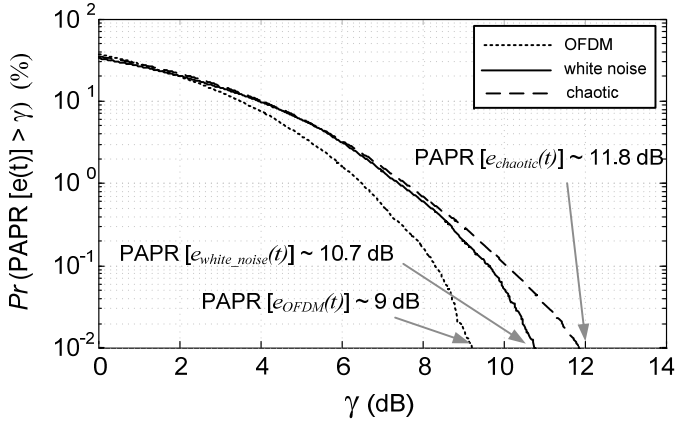
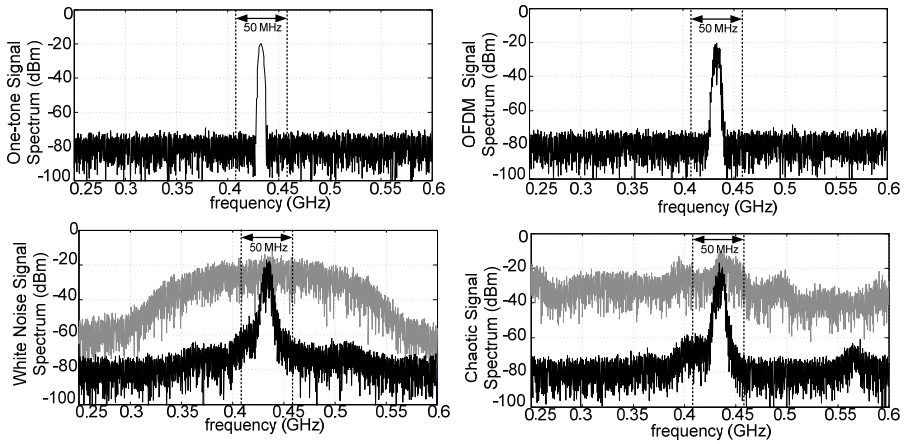
Signal Design

Signal	PAPR (dB)
1-tone	3
OFDM	12
White noise	13.7
Chaotic	14.8



PAPR[x(t)] ~ PAPR[e(t)] + 3 dB





A. Collado, A. Georgiadis, 'Optimal Waveforms for Efficient Wireless Power Transmission,' IEEE Microwave and Wireless Components Letters, 2014, to appear.

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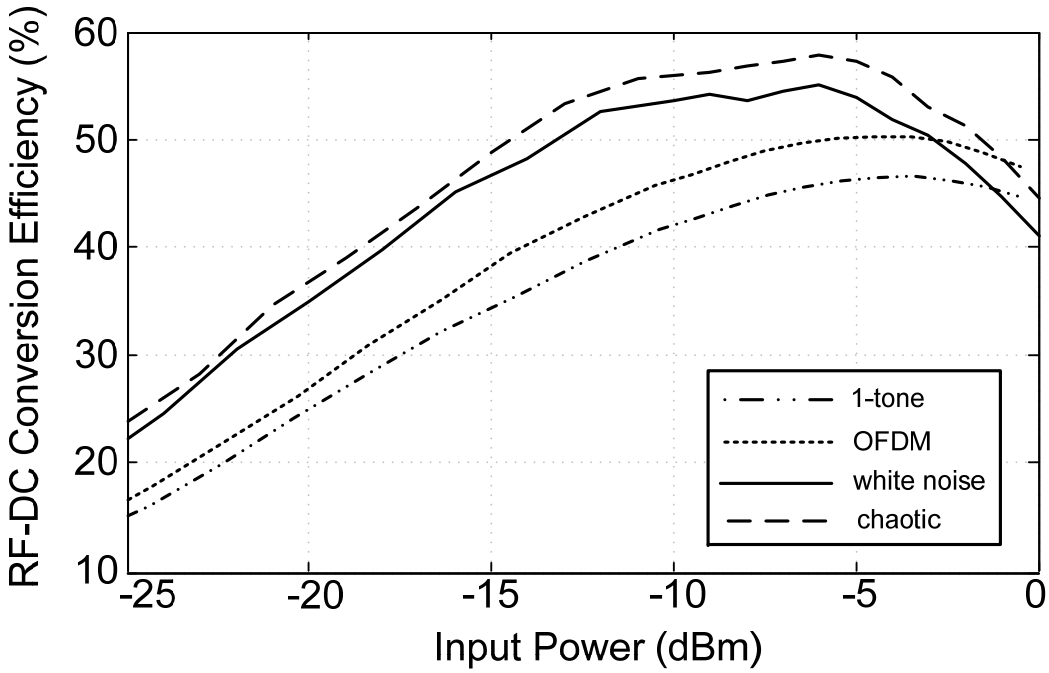
Signal Design

- rectifier operates at 433 MHz
- Skyworks SMS7630-02LF diode
- output load of 5.6 KOhm



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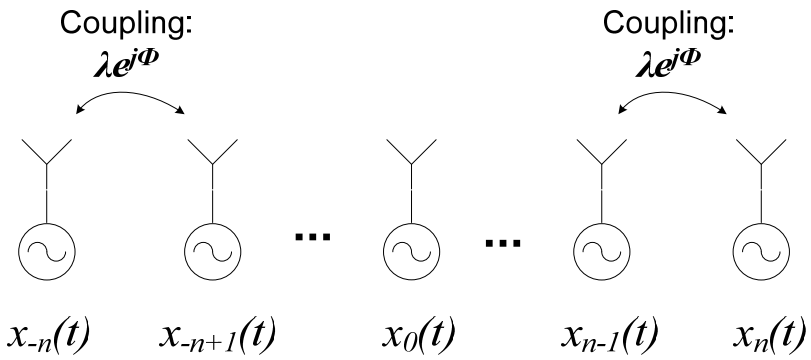
Signal Design

- High PAPR signals saturate the PAs
- Spatial power combining – each tone amplified independently and then combined in free space
- Mode-locked coupled oscillators – establish phase reference and control phase shift among elements



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A. Boaventura, A. Collado, A. Georgiadis, N.B. Carvalho, 'Spatial Power Combining of Multi-sine Signals for Wireless Power Transmission Applications,' IEEE Transactions on Microwave Theory and Techniques, Special Issue on Wireless Power Transfer, 2014, accepted for publication

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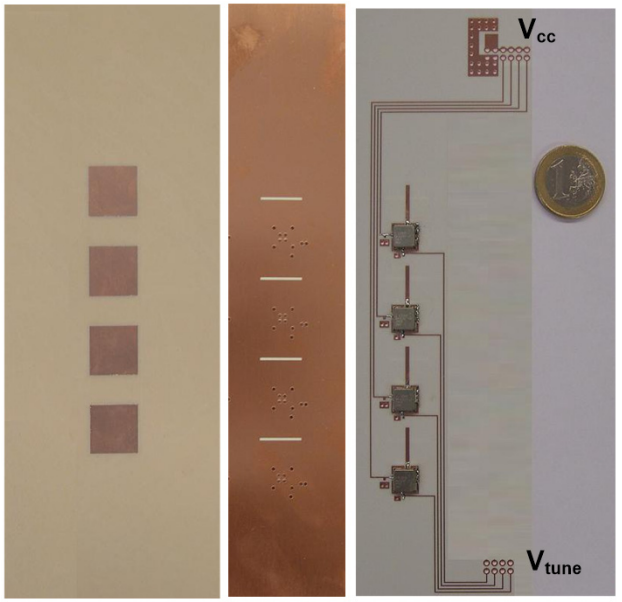
Signal Design

- 4x1 active antenna oscillator array at 6 GHz
- Patch antenna aperture coupled to a VCO



A. Boaventura, A. Collado, A. Georgiadis, N.B. Carvalho, 'Spatial Power Combining of Multi-sine Signals for Wireless Power Transmission Applications,' IEEE Transactions on Microwave Theory and Techniques, Special Issue on Wireless Power Transfer, 2014, accepted for publication

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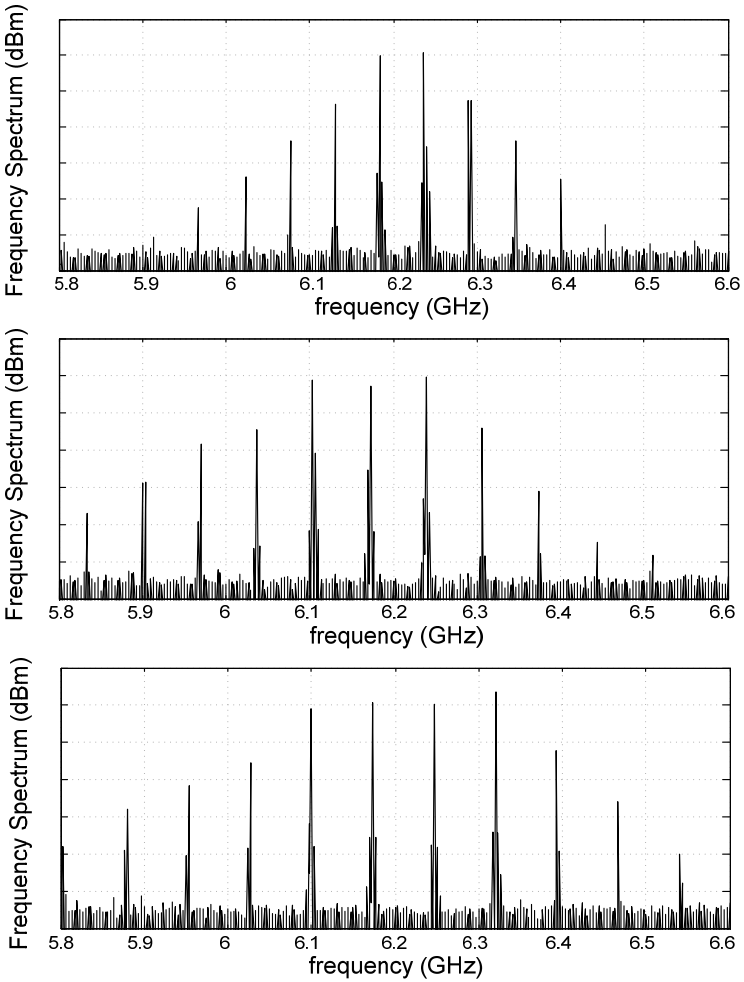
Signal Design

- Step1: 2 VCOs with 50 MHz spacing. Mixing products are created
- Step2: 3 VCOs. The third one with a free running frequency corresponding to one of the mixing products
- Step3: 4 VCOs. The fourth one with a free running frequency corresponding to one of the mixing products



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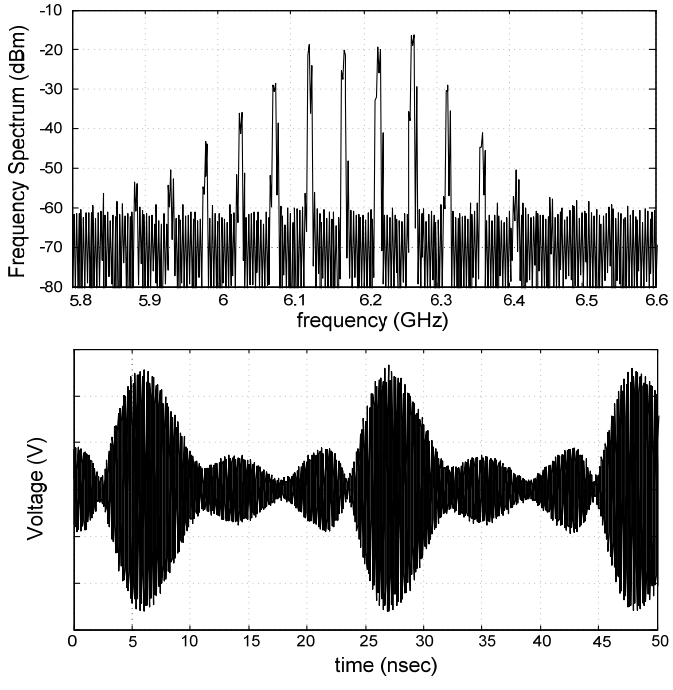
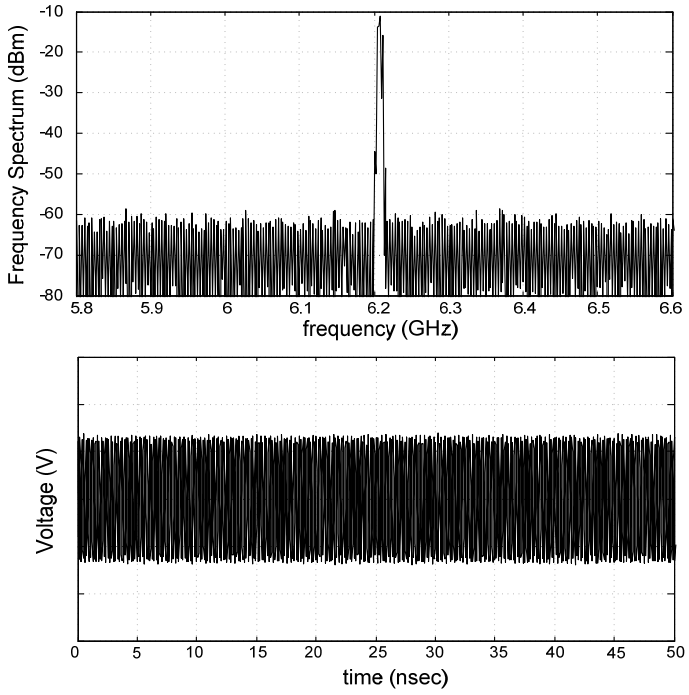
Signal Design

- Comparison of obtained DC voltage by a rectifier when using:
  - generated mode-locked signal with high PAPR signal
  - single carrier signal
- Same total average power for both signals

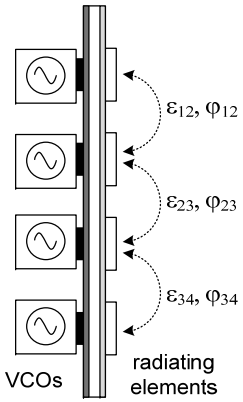


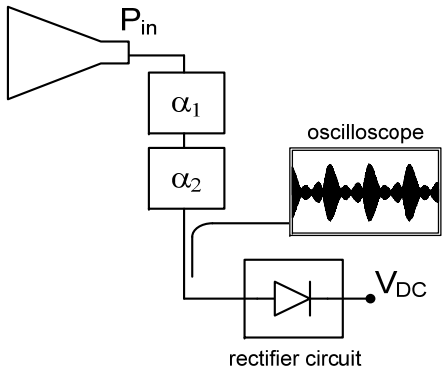


A. Boaventura, A. Collado, A. Georgiadis, N.B. Carvalho, 'Spatial Power Combining of Multi-sine Signals for Wireless Power Transmission Applications,' IEEE Transactions on Microwave Theory and Techniques, Special Issue on Wireless Power Transfer, 2014, accepted for publication

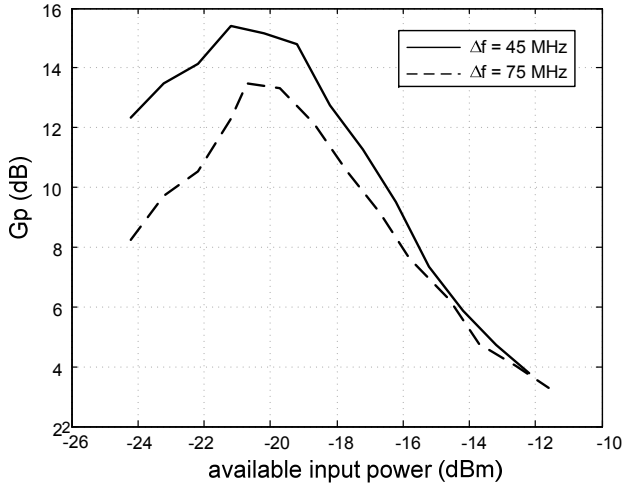
Signal Design

- Power gain compares the obtained DC voltage by a rectifier when using the high PAPR signal in comparison with a one-tone signal
- Improved performance when using the high PAPR mode-locked signal





$$G_{P1}(dB) = 10 \cdot log10\left(\frac{P_{DC\,(N)}}{P_{DC\,(1)}}\right) = 10 \cdot log10\left(\frac{V^2_{DC\,(N)}}{V^2_{DC\,(1)}}\right)$$



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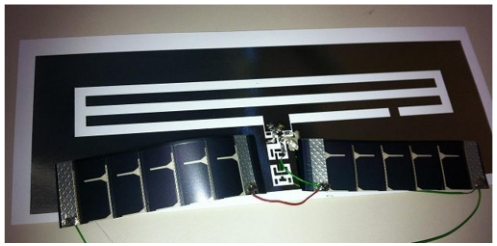
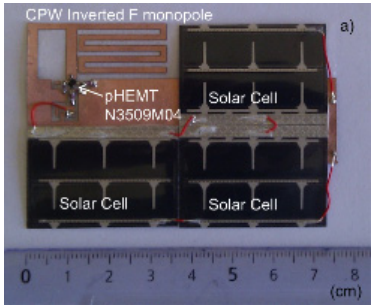
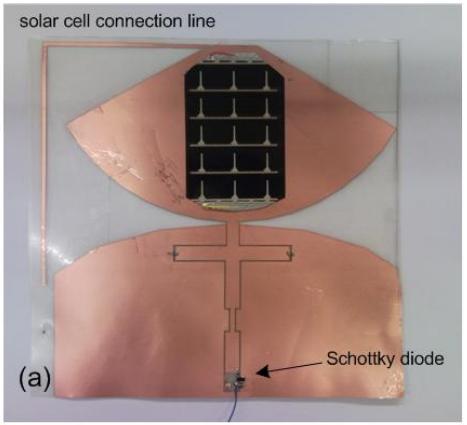
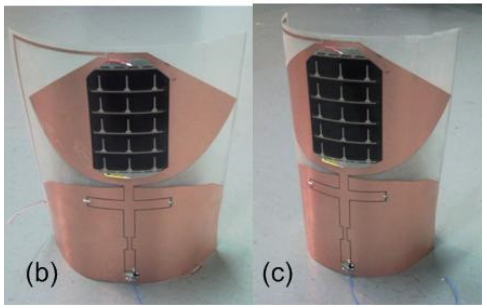
Challenges - Applications

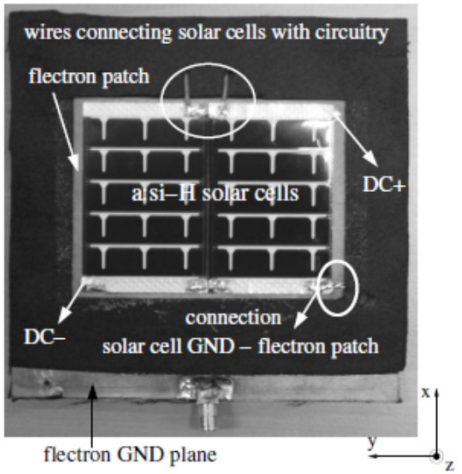
- **Multi-technology harvesters**
  - Solar antennas and rectennas
- **Flexible electronics**
  - Paper - Textile – Plastic substrates
- **Solar powered batteryless circuits**

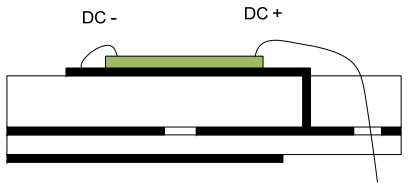












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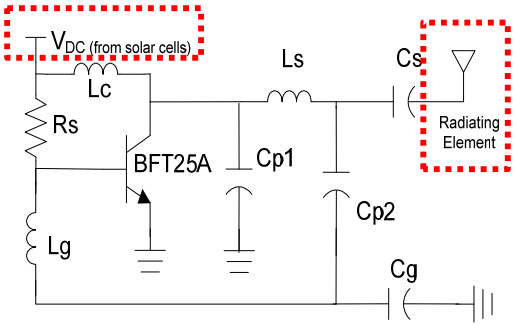
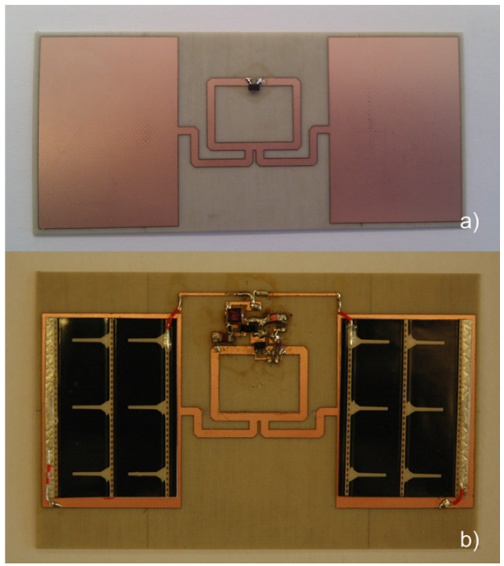
Challenges - Applications

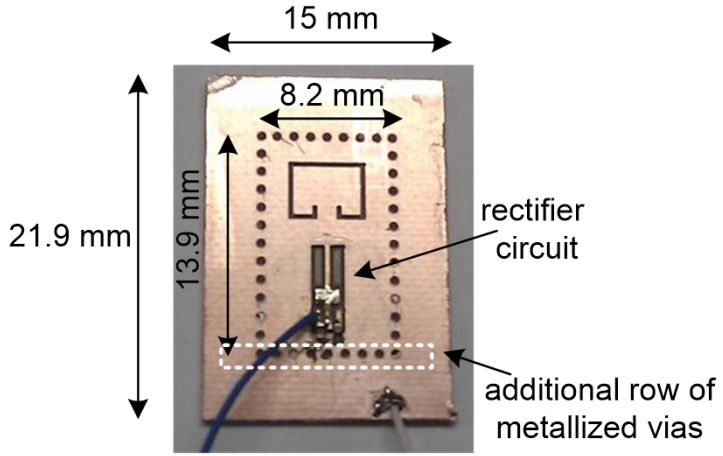
- **Solar RFID tag**
  - Solar => DC => RF
  - UHF Class-E oscillator
  - Solar antenna

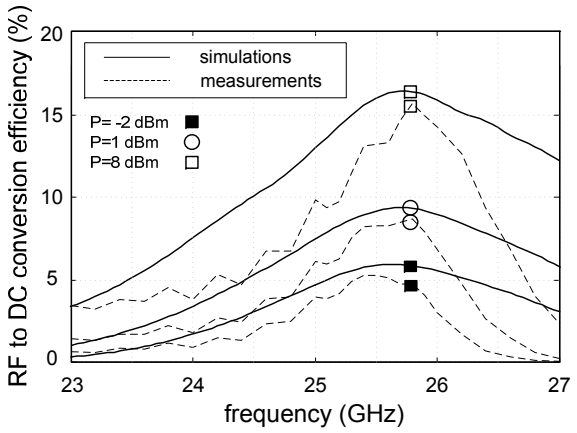




•A. Georgiadis and A. Collado, "Improving Range of Passive RFID Tags Utilizing Energy Harvesting •and High Efficiency Class-E Oscillators," in Proc. EuCAP 2012, Prague, 26-30 March 2012.

- **MMID (SIW)**
  - SIW 24 GHz rectenna
  - Integrate rectifier inside SIW cavity





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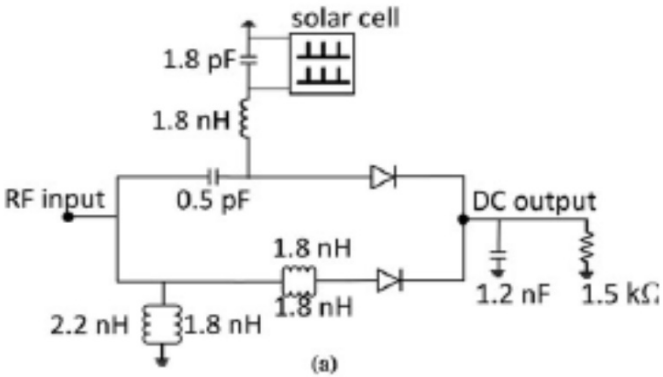
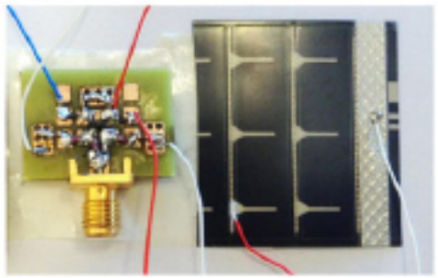
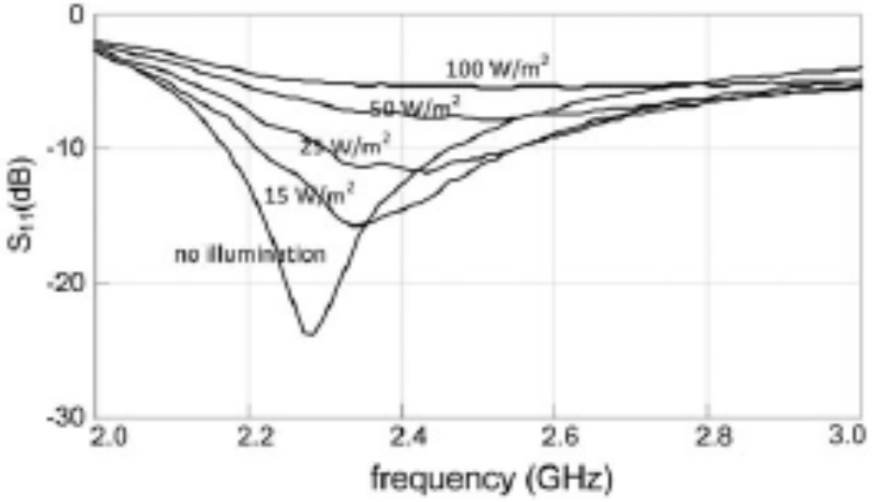
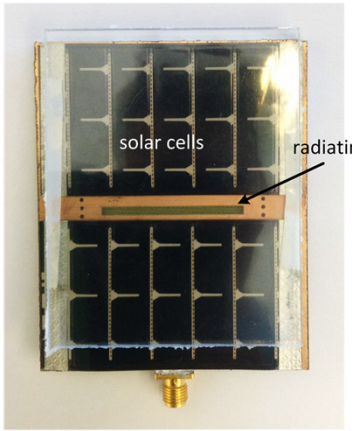
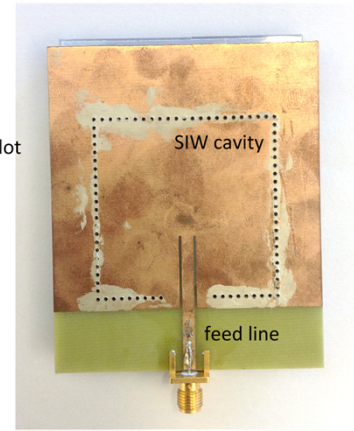
Multiple Technology Harvesters
Solar / EM

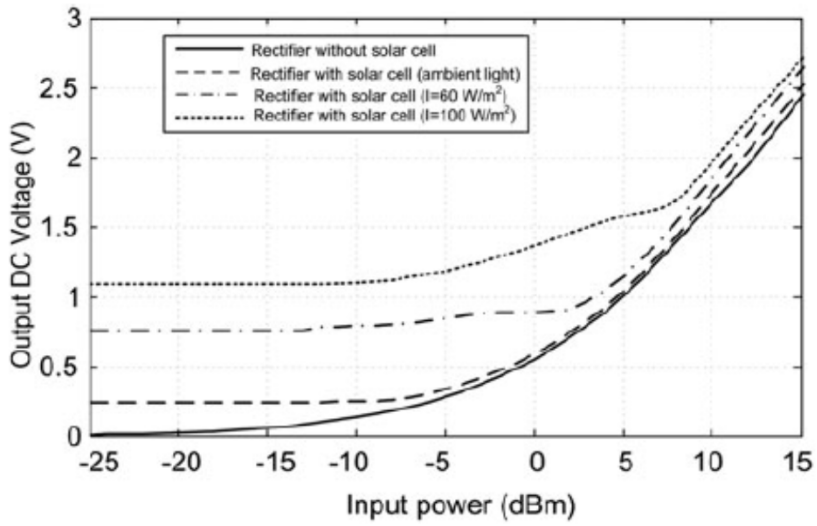






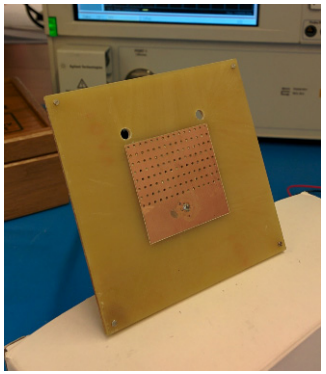


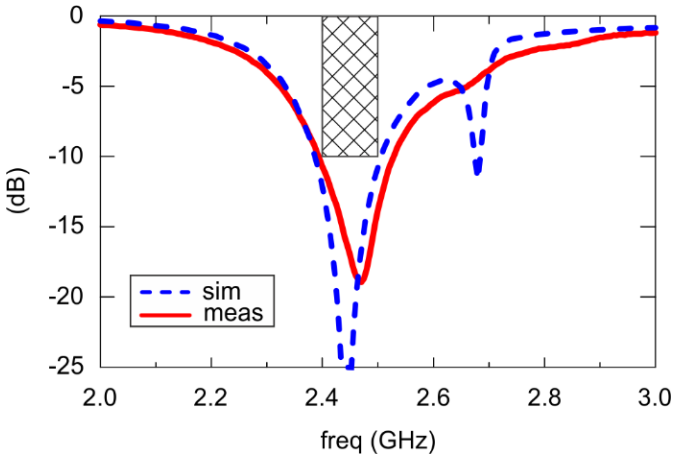


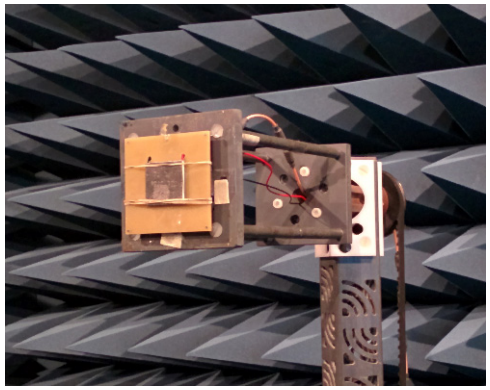


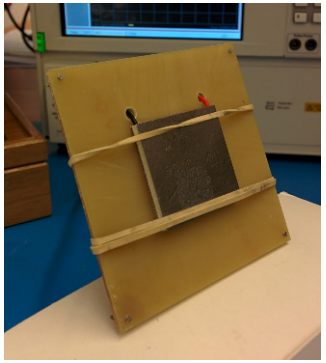
K. Niotaki, F. Giuppi, A. Georgiadis and A. Collado. Solar/EM energy harvester for autonomous operation of a monitoring sensor platform . Wireless Power Transfer, vol. 1, no. 1, pp. 44-50, Mar 2014.

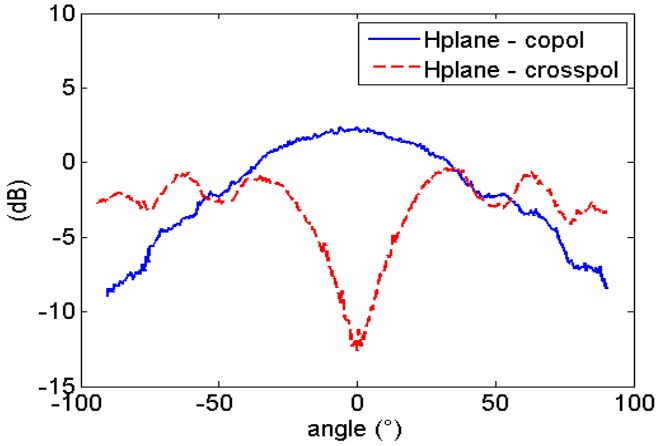
Multiple Technology Harvesters
Thermal / EM

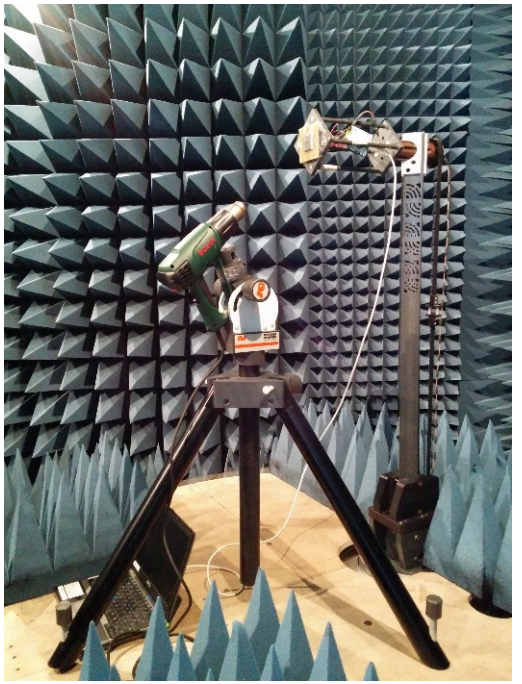












M. Virili, A. Georgiadis, K. Niotaki, A. Collado, F. Alimenti, P. Mezzanotte,
L. Roselli, N.B. Carvalho, 'Design and Optimization of an Antenna with Thermo-Electric Generator (TEG)
for Autonomous Wireless Nodes,' in Proc. 2014 IEEE RFID-TA, Tampere, Finland, 8-9 Sep 2014

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Conclusion

- Multi-band rectennas allow wider application
- Reactive networks capable of minimizing rectenna efficiency sensitivity to load variation
- High PAPR leads to higher efficiency
- Spatial power combining for WPT transmitters

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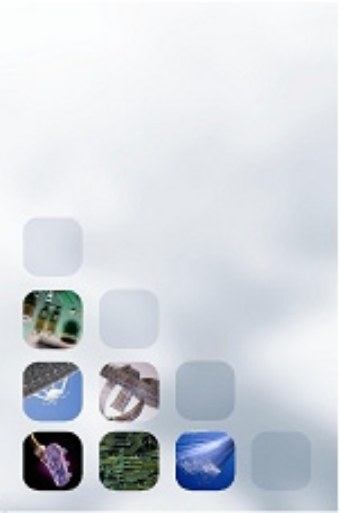

Cambridge Journal on Wireless Power Transfer

http://journals.cambridge.org/action/displayJournal?jid=WPT

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Vol. 1. No. 1. March 2014

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2014 IEEE RFID-TA Conference
8-9 Sep 2014, Tampere, Finland

http://www.rfid-ta2014.fi/

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This is the authors'version of the paper that has been accepted for publication in IEEE Potentials Magazine

A Rethinking of RF Wireless Power Transfer: How to Re-Green the Future Networks?

Ha-Vu Tran[1] and Georges Kaddoum[1]

arXiv:1711.01320v1 [cs.IT] 3 Nov 2017

***Abstract*—Green radio communication is an emerging topic since the overall footprint of information and communication technology (ICT) services is predicted to triple between 2007 and 2020. Given this research line, energy harvesting (EH) and wireless power transfer (WPT) networks can be evaluated as promising approaches. In this paper, an overview of recent trends for future green networks on the platforms of EH and WPT is provided. By rethinking the application of radio frequency (RF)-WPT, a new concept, namely *green* RF-WTP, is introduced. Accordingly, opening challenges and promising combinations among current technologies, such as small-cell, millimeter (mm)-wave, and Internet of Things (IoT) networks, are discussed in details to seek solutions for the question with how to re-green the future networks?**

I. Introduction

Over past few years, green radio communication has drawn much attention from the research community, and it has strong impacts on various aspects, such as telecom business, wireless technologies, and natural environments. Specifically, the electricity cost and CO_2 emissions have been increasing due to wireless network operation. For instance, the number of base stations (BSs) is more than 4 million, and each BS consumes an average of 25MWh per year (estimated approximate 80 percent of the total network's power consumption). Bearing in mind the environmental perspective, generating sufficient power to supply the networks causes a significant amount of CO_2 footprint. Particularly, the overall footprint of information and communication technology (ICT) services, e.g., computer, cell phone, and satellite networks, is predicted to triple by 2020 [1].

Recently, towards a future green world, energy harvesting (EH) technique has the potential to deal with the problem of energy inefficiency [2]-[5]. The main advantages of this approach can be presented in two-fold. First, the EH techniques harnesses green energy from natural sources, e.g., solar and wind. Thus, it contributes to reducing the overall footprint in order to protect surrounding environments. Nowadays, the popularity of using conventional energy sources, e.g., diesel, still dominates the use of the green sources. However, although the overall implementation cost of EH solutions is higher than that of conventional ones, this cost might be compensated after several years of operation. Second, another main challenge in future networks is prolonging the lifetime of smart user devices. Given this concern, EH networks take a tremendous advantage in various specific applications. For instance, EH is an efficient solution for reducing battery replacement costs in wireless sensor networks. Also, it can recharge the devices working in areas where the traditional power supply is infeasible, e.g., robotic devices working in toxic environments. Nevertheless, the amount of harvested energy from natural resources, such as solar and wind, may vary randomly over time and depend on locations and weather conditions. In other words, harvesting energy from these sources is not controllable and sustainable. For instance, there exists insufficient sunlight at night to generate energy, or it is difficult for indoor devices to harvest solar energy. In this context, radio frequency (RF) wireless power transfer (WPT) might be a promising approach to overcome such a drawback.

In this paper, we provide a comprehensive review to address the mentioned question of regreening the future world. In more specific details, the main contributions of this paper can be summarized as follows. First, by rethinking the use of RF-WPT, a concept so-called *green* RF-WTP is introduced. Second, a vision of future green networks based on the platforms of EH and green RF-WPT is presented. Thus, we discuss potential scenarios with the purpose of bridging green resources to indoor energy-hungry devices in the networks. Third, applications of various interesting concepts, such as small-cell, mm-wave, internet of things (IoT), etc., networks are highlighted. Given this concern, the challenges on each concept are identified. Further, we discuss some attractive combinations of the existing concepts, such as a mixture of full-duplex, RF-WPT, and mm-wave, and investigate how the latter works together properly. To this end, promising trends in future are drawn to provide solutions for future re-greened networks.

II. Energy Harvesting and Green RF Wireless Power Transfer

In this section, an overview of EH models and a discuss on green RF wireless power transfer are shown.

A. EH Models

EH methodology might be described by harnessing energy from surrounding environments or thermal and mechanical sources, and then converting the latter into electrical energy. The generated electrical current can be used to supply devices

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by RF wireless power transfer. Generally speaking, EH models can be classified into two architectures with harvest-use, and harvest-store-use. In the first one, energy is harvested, and then is used instantly. Besides, given the second one, energy is harvested as much as possible and then stored for future uses. In the following, we discuss the characteristics of such models.

In the harvest-use architecture, the EH systems directly supply devices. To guarantee the operation of the devices, the power output of the EH systems should be higher than the threshold of minimum working requirements. Otherwise, the devices would be disabled because there is not enough power supplied. As a consequence, unanticipated fluctuation in harvesting capacity close to threshold yields the working devices to vacillate in ON and OFF states.

Further, the harvest-store-use model includes a component storing harvested energy and also powering the connected devices. Thanks to the storage, the energy can be harvested until sufficient for supplying the devices. Moreover, such energy might be stored for later uses when there is lack of produced energy or the devices need to increase the working performance. The storage component might include two parts of primary and secondary storages. In this context, the secondary storage can be seen as a backup one. In particular, the harvest-store-use system can make non-stable but foreseeable energy sources, such as solar and wind, more favorable in uses.

B. Green RF Wireless Power Transfer

Over the last decade, solar, wind, mechanical, and thermal energy can be considered as the most efficient resources generating green energy usable for wireless networks. However, the main drawback of such sources is the lack of stability. In the quest for an alternative solution, the research community has explored that radio signals belonging to a frequency range from 300 GHz to 3 kHz can be used to carry energy over the air [3], [6], [7]. On this basis, a transmitter can proactively recharge wireless devices by sending energy-bearing RF signals whenever it is necessary. This is the principle of a so-called RF-WPT technique.

In fact, it is well-known that EH is a green technique since it helps to reduce the footprint. However, in a shared vision, the RF-WPT technique seems to be harmful to surrounding environments because it costs electricity to generate RF signals, and causes electromagnetic pollution to the human body as well as interferences to data transmission. By rethinking the role of RF-WTP technique, we suggest that the RF-WTP can be seen as green *iff* (i) the RF signal carrying energy is generated using power harvested from green resources (for example, BSs are connected with outdoor energy harvesters to harvest-and-store green energy and then use such energy to wirelessly recharge to indoor devices using RF signals), and (ii) a tight restriction is applied for increasing the transmit power (i.e., following the equivalent isotropically radiated power (EIRP) requirement approved by Federal Communications Commission). In this work, the RF-WTP satisfying such two conditions is so-called the *green* RF-WTP. The characteristics of the green RF-WPT technique can be listed as follows

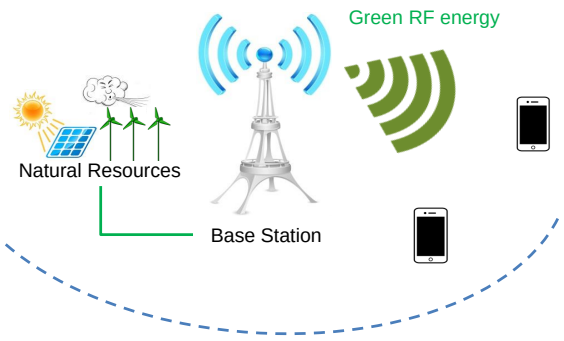

Fig. 1. Green RF-WPT.

- The green RF-WPT technique plays a role as a bridge between green energy sources and energy-hunger devices.
- The energy harvested at a receiver is foreseeable.
- The amount of harvested energy belongs to transmit power, propagation loss, and wavelength.

Indeed, it is expected that the green RF resource is one of the most interesting candidates for future applications.

III. A Vision of Future Green and EH Networks

A. A predicted model of future green networks with EH

Future networks, e.g., 5G, are expected to support multimedia applications to achieve 1000-fold higher throughput, 1000-fold higher mobile data per unit area, and 10-fold longer lifetime of devices over the fourth generation (4G) networks [4], [5]. To adapt this progress, the design of new cellular networks tends to a new form embracing a large-scale deployment of small-cells. Generally, the small-cells can be classified into distinct types including femtocell, microcell and picocell. Indeed, the multi-tier HetNet attains a promising gain in terms of spectral and energy efficiencies due to low power consumption and good ubiquitous connectivities [8].

On the other hand, nowadays, the development of wireless networks has broken the limits of power consumption, especially in cellular networks. Moreover, the energy cost and CO_2 emissions have been promptly growing due to the network operation. Specifically, this has inspired researchers with a challenging topic so-called future green wireless networks. As a promising solution, EH techniques exploit natural sources, and then contribute to reducing the overall footprint and extending the network lifetime. Nevertheless, the natural resources may not be always available to all devices. For instance, it is difficult for indoor devices to harvest solar energy. This yields another trend that BSs connected with outdoor harvesters which can harvest-and-store green energy when natural resources are available. Afterwards, BSs use such energy to wirelessly charge user devices using RF signals.

Another approach to future green networks is the concept of green IoT [10]. In fact, IoT is an emerging trend that billions of identified low-power devices, e.g., sensor nodes, are connected to each other without the need for human interaction. It can provide solutions to cut CO_2 emissions, reduce

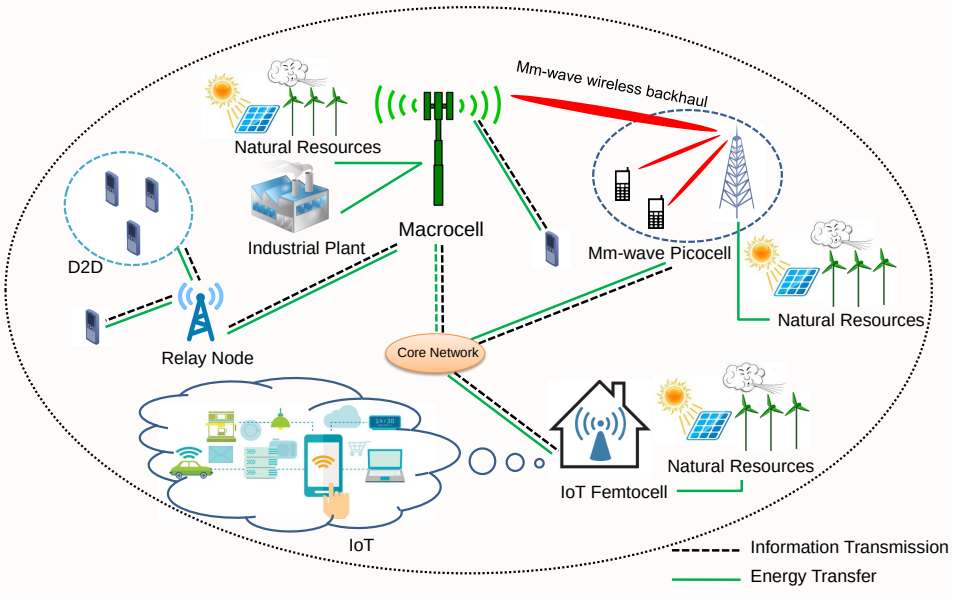

Fig. 2. A Future Green Network.

electromagnetic pollutions and improve energy efficiency. For instance, with tracking of motion sensors, the lights in rooms where there is no one inside would be turned off. Also, the green IoT technology can monitor energy usage in hi-tech buildings to reduce wasted energy. In fact, the green IoT is expected to enhance all technical, economical and environmental benefits. In particular, IoT network architectures mainly rely on the platforms of wireless sensor networks (WSNs), and cooperative networks to connect devices together [9]. In this concern, battery recharging for a large number of IoT devices is challenging indeed. Therefore, enabling the green IoT concept towards future networks requires advanced solutions of prolonging device's life-time, resource managements, and energy-efficient communication protocols.

Taking all the problems of interest into account, in the following, we further discuss several potential concepts towards green future networks as shown in Fig. 2. Specifically, challenges in implementing each concept are identified.

B. Green radio communications: Main concepts and discussions

1) Full-duplex networks: as mentioned, indoor devices which might not harvest green energy directly can be wirelessly powered by RF signals sent from BSs. Accordingly, this has inspired a combination of the full-duplex and simultaneous wireless information and power transfer (SWIPT) techniques. At the same time, the devices can receive energy in the downlink transmission while conveying information in the uplink connections to boost spectral efficiency.

Furthermore, we discuss two potential research issues of full-duplex SWIPT systems. First, the antennas at the full-duplex node are conventionally divided into transmit and receive sets. To improve the performance of SWIPT systems, an advanced form of the full-duplex technique that each antenna can simultaneously send information/energy and receive energy/information in the same frequency band is highly desirable. Indeed, this approach mainly depends on new hardware designs and new innovation of self-interference cancellation techniques. Second, full-duplex SWIPT small-cell base station can provide a promising approach regarding wireless backhauls in HetNets. In downlinks, the small-cell BS can receive information from macro-cell and transmit information/energy to users simultaneously. In uplinks, the small-cell BS can receive information/energy and send information to macro-cell at the same time. On this basis, the small-cell BSs do not require a separate frequency band of backhaul connections. Hence, resources and implementing costs are reduced.

2) Millimeter-wave networks: benefitting from mm-wave transmission, the overall electromagnetic field (EMF) exposure and power consumption per bit transmitted of networks are reduced due to a higher free-space attenuation at the mm-wave frequency, high directive antennas, and short distance links. Therefore, mm-wave communication has been considered as a primary candidate for green cellular networks in future. This approach is expected to achieve multi-gigabit data rates due to large spectrum resources at a ultra-high frequency band. Specifically, EH devices can extract energy from incident RF signals. Moreover, in mm-wave systems, many BSs are densely deployed to ensure proper coverage to ultra-high frequency networks. Thus, this can be attractive for the EH devices to potentially harvest sufficient energy.

In ultra-high frequency bands, the mm-wave signals mainly suffer from the propagation loss, such as poor penetration and diffraction. To make mm-wave networks more favorable for SWIPT, beamforming techniques can be a promising solution to increase the network coverage and system performance. Moreover, in mm-wave networks, although the small wavelength signals allow large antenna arrays to offer high beamforming gains, they require the alignment between transmit and receive beams to reach the highest possible

performance. In these concerns, there is a non-trivial problem of the beamwidth design. In practice, the length of beam-searching overhead is directly proportional to the number of beamformer candidates. The narrower beamwidth is, the more number of the beamformer candidates is, and then the longer overhead is. This leads the time of data transmission to be decreased. In constrast, the wider beamwidth means that it is easier for transmit and receive beams to be aligned, and the beam-searching process is sped up, however, the beamforming gain is reduced. Therefore, future works should consider the impact of beamwidth in various contexts to maximize the SWIPT system performance.

3) Wireless sensor networks: over the last few years, the trend for WSNs is one of the most attractive topics due to flexible installation and convenient maintenance. Accordingly, many standards such as WirelessHART, WIA-PA, and ISA100.11a have been proposed. Particularly, the IoT technology is mainly implemented on the WSN platform. With an integration of IoT with sensors, sensor devices can be interconnected with the global Internet in order to provide solutions for future networks, such as reducing wasted energy in hi-tech buidings.

Specifically, replacing or charging the batteries in IoT WSNs may take time and costs due to a large number of sensors, and this process becomes dangerous for humans in hazardous environments. As a result, the EH from natural resources and RF signals for WNSs have been considered as a promising solution to prolong the sensor's lifetime.

As the main distinction from conventional WSNs, EH-WSNs require new criteria in the fairness between information transfer and EH requirements. In fact, the network can fail to adapt the EH requirement while ensuring other system performances, such as throughput, delay or packet loss. As a result, leveraging between data transmission and EH is one of the critical concerns in designing EH-WSNs. Therefore, efficient resource allocation schemes should take such a problem into account to achieve high energy efficiency for EH-WSNs.

4) Cooperative relay networks: in recent, the cooperative relay network have been evaluated as one of the main core networks for the IoT technology where IoT nodes can communicate with each other and forward information and energy to the remote nodes. Up to now, many mature research works of cooperative communication have clearly shown that the relay can be implemented not only to extend the coverage range but also to improve the performance of wireless communications. Further, the concept of EH/WPT relay networks has been proposed and studied to enhance the lifetime of devices and overall performance of wireless networks. In cooperative EH/WPT relay networks, improving performance gain on the physical layer is one of the main research directions. Hence, most previous efforts attempt to design novel schemes regarding relay operation policies, power allocation, precoder optimization and relay selection.

Considering existing challenges, the enhancement for both spectral and energy efficiencies in cooperative relay networks with the green RF-WTP is remarkable. In this concern, full-duplex or two-way relaying methods might be a promising solution. Further, it is suggested that developed resource allocation schemes should consider the influence of incomplete channel state information (CSI) (e.g., the relay nodes have a partial user's CSI), and the energy status at the relay nodes and users (e.g., the available energy, current power consumption, predicted energy harvested from natural resources or RF signals, etc.) on system performance.

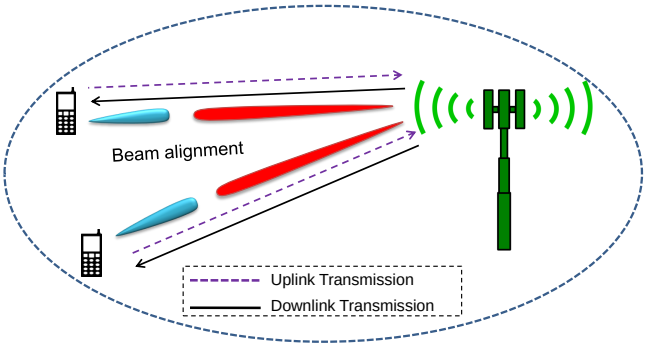

Fig. 3. Mm-wave SWIPT networks with full-duplex communications.

IV. Future Research Issues

In the prior section, several challenges of each concept have been presented. In future networks, since wireless communication systems are expected to be a mixture of various novel system concepts to enhance both the spectral and energy efficiencies, some interesting combinations of the existing concepts are discussed as follows.

A. When full-duplex communications meet mm-wave SWIPT networks?

A combination of mm-wave and full-duplex SWIPT systems, as shown in Fig. 3, seems to be interesting. Most of the recent research on full-duplex SWIPT systems mainly address the communications in conventional frequency bands. However, in mm-wave frequency bands, there are two main challenges need to be discussed. First, the practical implementation of mm-wave full-duplex SWIPT should be investigated with a bandwidth of several GHz. Second, in mm-wave networks, the communication is inherently directional. Therefore, at both the transmission and reception sides, the directional antenna should be used. As a result, the node structure should be re-considered according to the characteristics of mm-wave signals. To reduce the cost of antennas (i.e. about parabolic antennas), one of the efficient solutions is to employ transmit and receive beams to limit self-interference. Therefore, the future objectives should address:

- studying how the beamwidth affects the beam alignment, beamforming gain and beam-searching process,
- investigating the performance trade-off between self-interference and data transmission time,
- and then properly allocating available resources to optimize system performance.

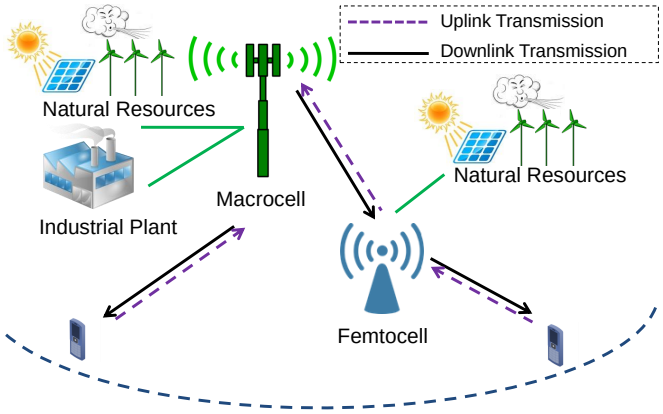

Fig. 4. SWIPT and EH HetNets with the full-duplex technique.

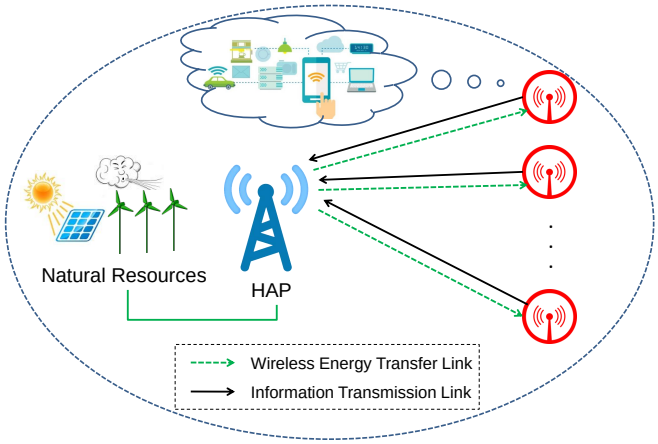

Fig. 5. RF-WPT Internet of Things networks.

B. What are potential scenarios for SWIPT and EH HetNets with the full-duplex technique?

Full-duplex EH-SWIPT HetNets may bring a bright, however, challenging approach. Given this concern, in Fig. 4, a macro-cell BS harvests energy from natural sources and then communicates with a small-cell BS. On the other hand, whereas the small-cell BS receives information from the macro-cell, it transmits information/energy to the users at the same time. In another case, the small-cell BS transmits information to the macro-cell while it receives information from the user simultaneously. Given these scenarios, the sytem benefits from an enhanced spectral efficiency, however, there appear a lot of interference sources, e.g., self-, inter-cell and intra-cell interferences due to full-duplex communications. Specifically, dealing with downlink-to-uplink interference is a big challenge since the downlink power dominates the uplink one in general. Therefore, using optimization frameworks, future works need to focus on:

- designing new self-interference cancellation techniques,
- managing the downlink-to-uplink interference,
- allocating resources to optimize the system performance in terms of the information and power transfer.

C. What are the main concerns of wirelessly powering Internet of Things networks?

Due to large-scale deployments of IoT networks, replacing or recharging the device's battery is one of the main challenges. Specifically, a large number of IoT sensors are implemented in indoor locations where natural resources might be not available to harvest directly. In this context, the green RF-WPT is a promising candidate for prolonging the lifetime of the IoT low-power devices. Thus, this implies a scenario as illustrated in Fig. 5 where a sink node is responsible for harvesting energy from natural resources and then wirelessly transferring power to devices in a IoT wireless sensor network. In addition, the devices can communicate with each other. Given the model, some important concerns need to be addressed as follows.

- scheduling the power transfer to energy-hungry users according to the harvested energy at the sink node,
- exploiting interference from ambient environments to improve the EH performance,
- maximizing the information performance while satisfying EH requirements.

V. Concluding Remarks

In this work, we have presented a review of promising trends towards future green networks. Based on the platform of EH techniques, several potential concepts such as HetNet, mm-wave and IoT networks, have been presented. In particular, we discuss a promising architecture, so-called green RF-WPT. In fact, the latter plays a crucial role as a bridge between natural energy resources and smart energy-hungry devices. Accordingly, we have shown a vision of future green networks in which smart devices can be recharged by green resources even when they cannot harvest energy directly. Furthermore, to facilitate the regreening process while adopting intensive system performance required in future networks, the combinations of techniques, such as SWIPT, mm-wave, full-duplex, can bring outstanding outcomes. Given this concern, we have identified some challenges in mixing the potential concepts, and discussed how they can work together. Indeed, it is expected that the green RF-WPT-based approaches can be one of the potential solutions for regreening the future ICT world.

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ENERGY AND RADIO SCIENCE

Cross dipole rectennas for satellite health monitoring

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Keywords: energy harvesting and wireless power transfer, rectennas, printed dipole array, satellite application

Abstract

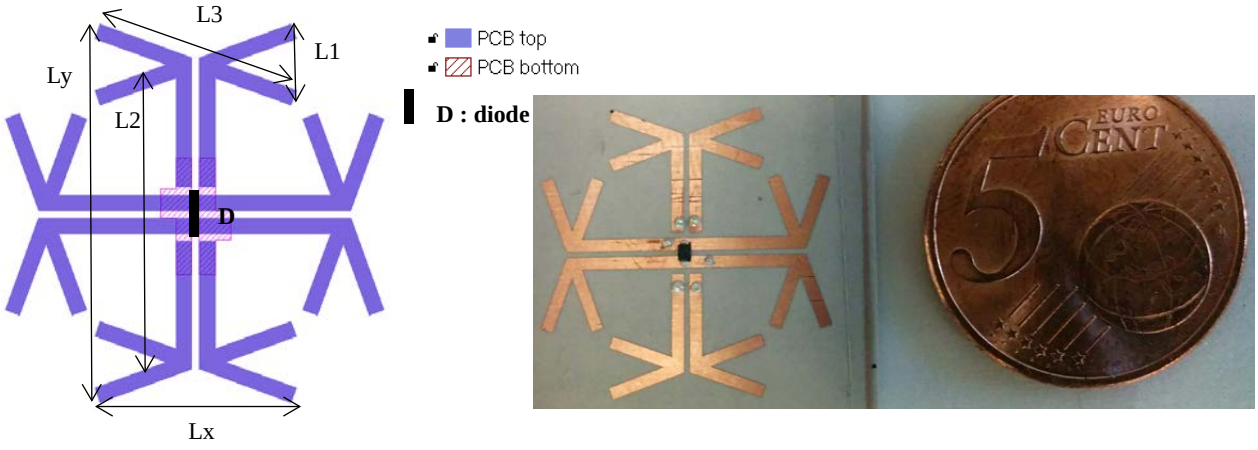
This paper addresses recent results obtained by using cross dipole rectennas designed for Ku band. The targeted application is the harvesting of the spill-over losses of microwave antennas in order to power autonomous wireless sensors used for satellite health monitoring.

Introduction

Recently, the harvesting of the spill-over losses of the microwave antennas was identified as a realistic solution to power autonomous wireless sensors for health monitoring of the antenna panels of the broadcasting satellites [1]. The electromagnetic power radiated by microwave antennas is almost constant and consequently the DC power regulatory circuits should be minimal for the harvesting systems/rectennas. Nevertheless the rectenna should be compact, energetically efficient and robust. Moreover it should be as simple as possible and compatible with space design requirements in terms of thermal, mechanical and radiation safety/immunity behavior. This paper presents recent results obtained in Ku band by using an innovative rectenna topology based on a compact printed cross dipole antenna array and low cost Schottky diode.

1. Topology

The proposed topology is composed by: four cross dipoles array (4CDAA) and a silicon Schottky diode – Skyworks SMS7630 [2]. The PCB layout and a photo of the manufactured rectenna is shown in Fig. 1. The antenna array and the diode are mounted on the top of the PCB while the RF shunt capacitor is mounted on the bottom side of PCB. A reflector metallic plate is added on the bottom side in order to improve the antenna gain. This compact topology (its size is fixed by the size of the antenna array) is an extension of a rectenna using two cross dipole antenna array 2CDAA presented in [3] and a non-resonant matching technique is used. This non-resonant matching is obtained by properly controlling (i) the input impedance of the antenna array and (ii) the input impedance of the rectifier itself (composed by the diode and the low pass RC filter implemented by using the shunt capacitor and the load).

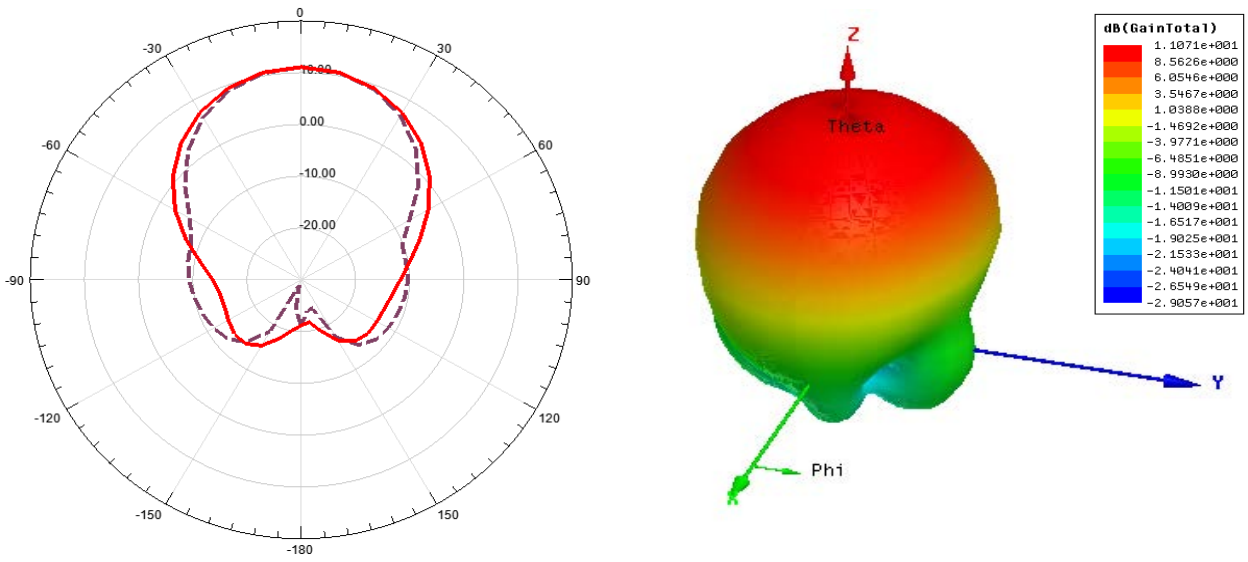


(a) (b)

Fig. 1. (a) Top view (not to scale) of the layout of the 4CDAA rectenna its main geometrical dimensions and (b) a photo of the manufactured rectenna

2. Design and Results

First, the cross dipole antenna was designed and optimized under HFSS software. Six via holes were used (as shown in Fig. 1) in order to connect the top and the bottom part of the PCB. The bottom side of the PCB (used to interconnect the cross dipole antennas and for the mounting of the RF shunt capacitor) was taken into account in the HFSS simulation model. The designed antenna was not measured separately. As shown in Fig. 1, the appropriate feed technique for such an array is based on the use of coplanar striplines. Consequently a complex transition between the coplanar stripline and a coaxial connector has to be designed and implemented before measuring such an antenna in a standard anechoic chamber. The simulated gain (maximum gain: 11dBi) obtained for the 4CDAA at the targeted operating frequency (12 GHz) as well as the 3D radiation pattern are shown in Fig. 2.



a) **b)**

Fig. 2. Simulated gain (HFSS) in the xOz plane (phi=0°, red continous line) and in the xOy plane (phi=90°, black dashed-line of the 4CDAA) at the frequency of 12 GHz.

Second, a generic electrical diode model existing in the ADS software was customized with the following parameters: ohmic resistance R_s = 20Ω, junction capacitance C_{j0} = 0.14 pF, saturation current I_s = 5 µA, forward voltage V_f= 0.34V, reverse breakdown voltage B_v = 2V and the current at reverse breakdown voltage I_{bv}= 0.1mA. This model allows performing an initial design of the PCB bottom side.

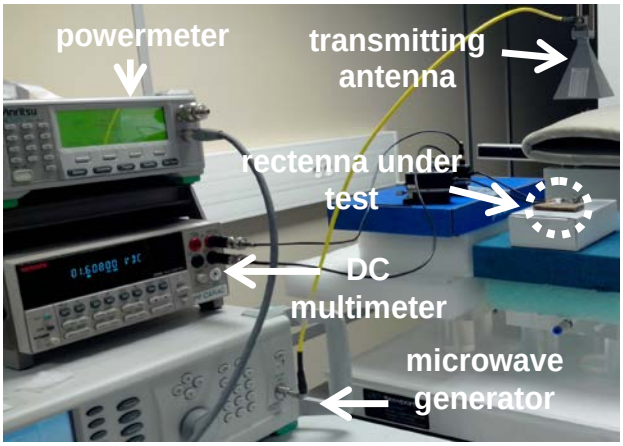
The experimental setup shown in the Fig. 3 was used to characterize the rectenna. A microwave signal generated from an Anritsu MG3694B generator was injected at the input of a horn antenna which illuminated the rectenna under test with a linear polarized E-field. An automatic acquisition routine was implemented in Labview software from National Instruments to speed-up the acquisition process. The harvested DC voltage was measured by using a DC multimeter. The DC power can be computed from the measured DC voltage as long as the load is known. The measured losses due to the coaxial cable and connectors between antenna and the signal generator were in the range of 3 dB in the operating frequency band. The main dimensions of the fabricated rectenna are: L1= 4 mm, L2= 15.3 mm, L3= 10.6 mm, Lx= 10 mm, Ly= 19.5 mm.
The efficiency η (in %) of the rectenna can be computed by using the following definition [4]:

$$\eta = \frac{P_{DC}}{S \cdot A_{eff}} \cdot 100 = \frac{4 \cdot \pi \cdot P_{DC}}{S \cdot G_R \cdot \lambda^2} \cdot 100 \qquad (1)$$

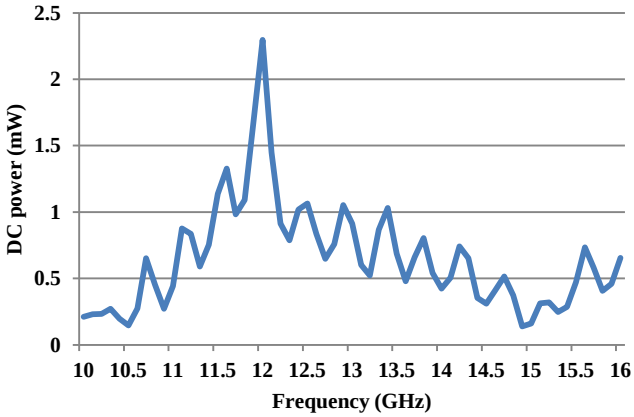
where P_{DC} is the harvested DC power, S is the incident electromagnetic power density, A_{eff} is the antenna effective area, G_R is the gain of the (rectenna's) antenna and λ is the wavelength of the illuminating electromagnetic wave. The power density (µW/cm^2) can be computed as a function of E-field effective value E (V/m) on the antenna surface or as a function of the RF power P_t injected to the input of the transmitting horn antenna of gain G_t and positioned at the distance *d* from the rectenna, as follows:

$$S = \frac{E^2}{120 \cdot \pi} \cdot 100 = \frac{30 \cdot P_t \cdot G_t}{d^2 \cdot 120 \cdot \pi} \cdot 100 \qquad (2)$$

The experimenal results obtained as function of frequency are reported in Fig. 2 for an optimal load RL=300Ω (this is the optimal load for 2CDAA rectenna). The following configuration was used : d=21 cm, G_t=19 dBi, power delivered by the microwave generator: 22 dBm and the injected power at the input of the transmitting horn antenna: 19 dBm. The simulated (by using HFSS software) gain of the receiving antenna (4CDAA) was G_R=11 dBi. The maximum DC power (2.3 mW) was obtained at 12 GHz for RL=300Ω. By using equation (1) and (2) we obtain : E~65.5 V/m, S~1138 μW/cm^2 and the efficiency of the rectenna is 32% at 12GHz.



(a)



(b)

Fig. 3. (a) Experimental setup used to characterize the rectenna and, (b) Measured DC power on a load of 300 Ω as function of frequency.

3. Conclusion

A rectenna based on the use of four cross dipoles antenna array was designed, manufactured and measured. The experimental results demonstrate that a DC power of 2.3 mW can be obtained if the rectenna is illuminated by an electromagnetic field of E=65.5 V/m. This DC power is sufficient to power a wireless sensor (including the sensing device, the power management unit and the wireless tranceiver) for satellite health monitoring application.

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A Linearly Polarized Rectenna for Far-Field Wireless Power Transfer

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ABSTRACT: A rectenna, which consist of a yagi Microstrip antenna along with rectifier circuit is proposed for wireless power transmission in the far field region. The wireless power link is examined between an insitable antenna and an external antenna. A method of adding directors in yagi antenna is used to increase the wireless power link, thus to enhance the efficiency of the designed antenna. After calculating an receiver power level of an insitable antenna, the rectifier circuit efficiency is optimized. Yagi uda antenna is operated at a frequency of 2.4GHz to 2.485GHz.

KEYWORDS: Far field wireless power transfer, insitable rectenna, ISM band, power conversion, Yagi uda antenna.

I. INTRODUCTION

Wireless technology have gained a lot of attention recently. Wireless power transfer can be attained by two methods namely near field and far field region. When compared to near field wireless power transmission, far field wireless power transfer has achieved greater advantages such as longer distances and accurate alignment avoidance. An insitable antenna has many challenges when designing for various requirements. The requirements such as radiation efficiency, acceptable bandwidth and compact size.

Many research have been conducted on designing implantable antennas[2]-[13]. In [3] and[4] designing of insitable antenna in different environment is provided. To improve and extend the lifetime of insitable devices, a dual band antenna was studied in [6]. In [11] and [12] for biomedical application, antenna such as slot and microstrip antenna was studied respectively.In other words for far field wireless power transmission an insitable antenna is a important component. An insitable rectenna is used to receive and store energy from RF power and it consists of an rectifier circuit and an implantable for wireless power transmission , an efficient rectenna is needed. To optimize the antenna design optimal diode RF,dc impedances and function of input power were obtained in[15]. Circuarily polarized rectennas were studied in[16]and[18]. A low input RF power 2.45GHz rectenna were designed in [19]. Rectifier circuit conversion effciency of 15.7% at -20 dBm is obtained in[21]. In [22], separate transmit and receive on-chip antenna was designed for pressure monitoring application. A triple band implantable antenna was designed for wake up controller at 2.45GHz in[23],[24]. The proposed antenna provide an efficiency of 78% at a frequency range of 2.45GHz. In [25] wireless power transmission through dispersive tissue was studied. In the proposed work harmonic termination technique in a single voltage doubler rectifier is used to boost the DC output voltage and conversion efficiencies.

To achieve an longer distance, an insitable rectenna is designed in this paper, shown in Fig. 1. The designed insitable yagi antenna is utilized at the operating frequency of 2.45GHz for the wireless power link. After increasing the resonant frequency of the antenna in the implantable device, the wireless power is transferred. The power level received by the yagi antenna is calculated based on the certain rules. A rectifier circuit is designed finally and efficiency of rectenna is estimated.

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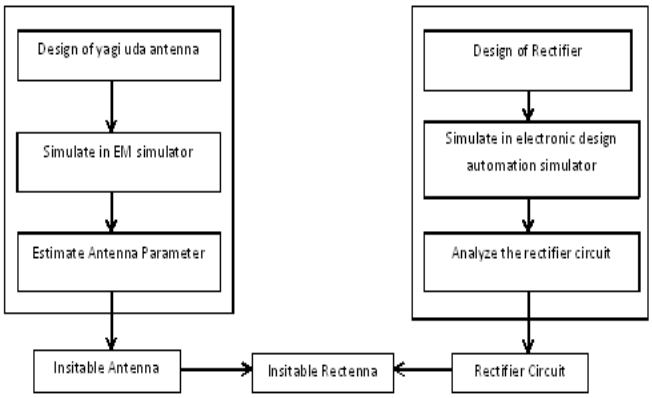

Fig. 1. Design of Yagi Uda antenna and rectifier circuit

II. INSITABLE ANTENNA DESIGN

A. Antenna Composition

An Insitable antenna shown in Fig. 2 is designed on theThree-layer FR-4 susbstrate with dielectric constant( ) of 4.4. The ground plane of yagi antenna are 100mm x 35mm. The dipole element height is 60.96mm and width is 3.05mm. Directors of yagi antenna are 3.4mm x 34.49mm.The spacing between dipole and directors are less than 0.3 .The reflector is designed greater then dipole and dipole is designed greater than directors. 2.4GHz were used in the simulation model. For design and analysis, Ansoft High Frequency Structure Simulator is used. Simulated reflection coefficient for the proposed antenna is shown in the Fig. 3. From Fig. 3. It is observed that the antenna covers from 2.3556GHz to 2.566GHz for $|S_{11}|$ less than -10dB. The gain obtained in the simulation is 8.14dBi at 2.45GHz is shown in Fig. 4. By tuning the resonant length and ground plane, the yagi antenna can operate at 2.45GHz with accurate matching. The geometry parameters of yagi antenna is shown in Fig. 2. The value of VSWR is estimated between 1 to 2 and results is shown in Fig. 3. The polar pattern of directivity is shown in Fig. 5. The comparison of frequency with reflection coefficient is shown in Fig. 6. The real and imaginary part of yagi uda antenna is shown in Fig. 7. Maximum efficiency of antenna is shown in Fig. 8. The loss tangent of an yagi antenna is 0.02.

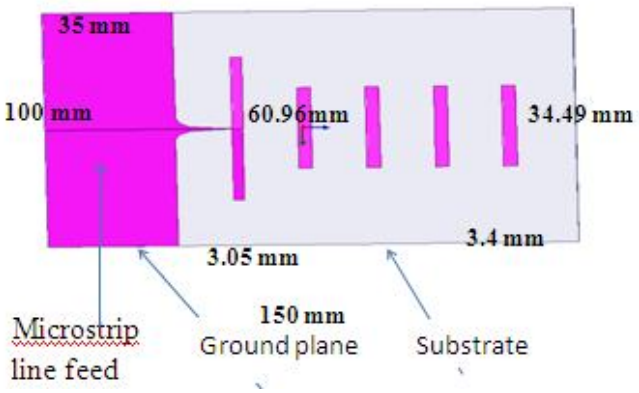

Fig. 2. Geometry of Yagi uda antenna

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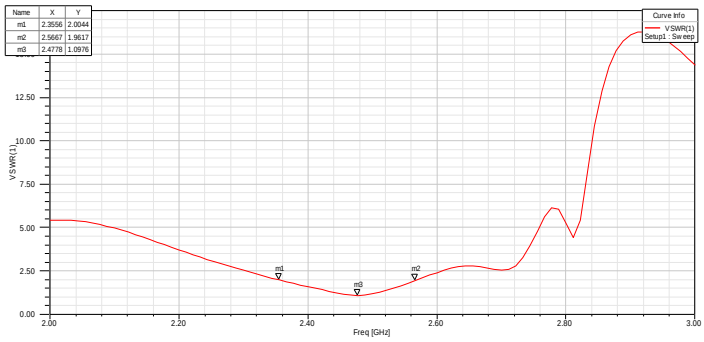

Fig. 3. Simulated VSWR of the Yagi antenna

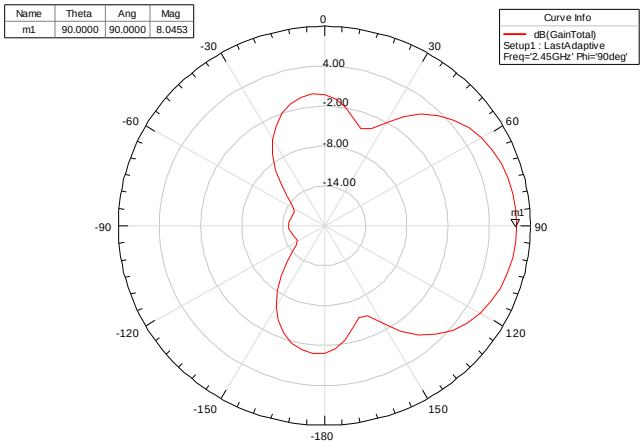

Fig. 4. Simulated Radiation pattern with gain(dBi)

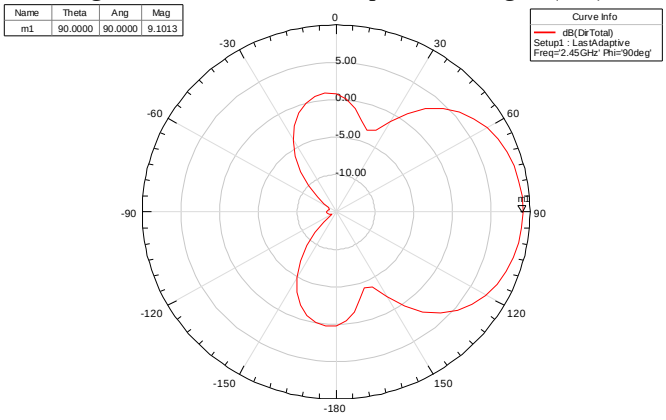

Fig.5. Simulated directivity(dB) with respect to theta

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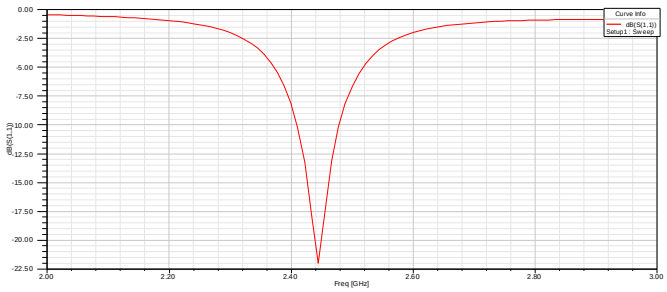

Fig. 6.Simulated Frequency Vs return loss

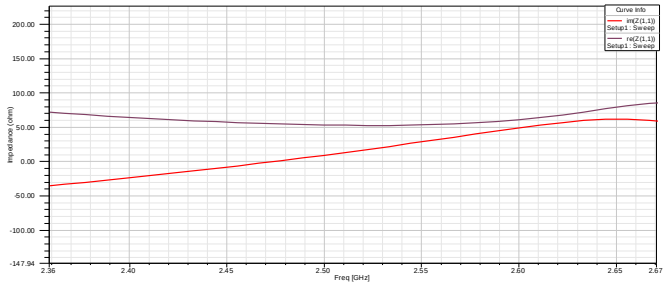

Fig. 7. Frequency Vs impedance

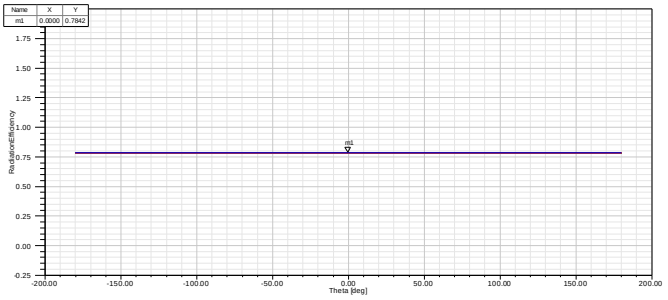

Fig. 8. Elevation angle Vs efficiency

B. Fabrication of Yagi Uda antenna

The designed antenna was fabricated on a copper material and substrate using standard photolithography technique. First the design was transferred on to a transparent mask. The copper material was cleaned, dried and applied using spinner. The thickness of FR4 substrate is 1.6mm. Above the substrate, an reflector with spacing of dipole and director is placed. Between the holder and reflector, an Microstrip line feed is soldered using copper material.One of the advantage of fabricating an yagi uda antenna is, the wider bandwidth is achieved. Fabricated yagi uda antenna is shown in Fig. 9.

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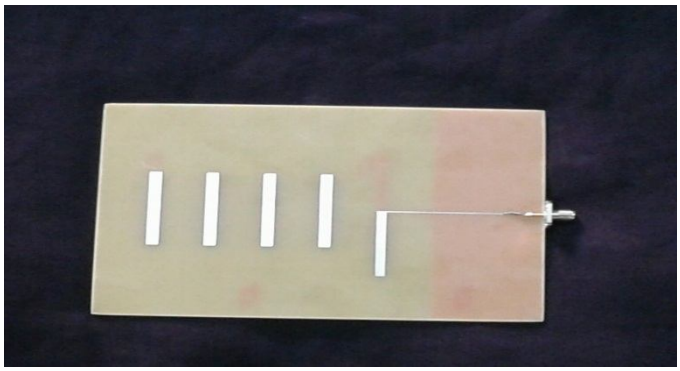

Fig. 9. 2.45GHz Fabricated Yagi Uda Antenna

III. ANTENNA FEEDING METHOD

The Microstrip antenna can be feed through various method. These methods is classified in to two types. 1. Contacting method, RF power is fed directly to the radiating antenna using a connecting element such as a Microstrip line feed. 2. Non-Contacting method, the EMF coupling is done directly to transfer the power between Microstrip line and the radiating element.

A. Microstrip Feeding method

The feeding method used in designing yagi uda antenna is Microstrip line feeding method. In this method, the conducting strip is connected from the reflector to the driven element is shown in Fig. 2. The strip which conducting the signal is vey thin, compare to the reflectors, driven elements and dipoles.

B. Measurement

By using homogeneous mixture liquid, the antenna was measured. 58.2% de-ionized water, 5.1% DGBE(Diethylene glycol butyl ether) are the recipes's used. The measured VSWR is shown in Fig. 3. Indicates that the measured impedance bandwidth of the proposed antenna is from 2.3556 to 2.557GHz for $|S_{11}|$ less than -10dB.

IV. RECTIFIER CIRCUIT

In this paper, the rectifier is proposed with schottky diode HSMS2860, SMA pF capacitor which is need to convert RF-to-DC for wireless power transfer.In order to improve the rectifier efficiency, the matching and filter is added between the antenna and rectifier and between the rectifier and the load.
HSMS2860 schottky diode was chosen because of its high speed and low voltage drop. In order to avoid impedance mismatching, the characteristic impedance is set to 50Ω with proposed insitable antenna. The proposed 2.45GHz linearly polarized Yagi Uda antennahas been designed and pertinent to the Microstrip rectifier to obtain the rectenna.

A. Fabrication of Rectifier

Matching stub, band pass filter, high frequency Schottky diode, DC filter, SMA capacitor is usedfor fabricating a rectifier. First we fabricate an ground plane and coated with copper material. Secondly, we placed an FR4 substrate on top of the ground plane. Thirdly, above the FR4 substrate, 5Ω resistor and capacitor is soldered, in between that schottky diode is soldered. Finally they are connected with transmission lines coated with copper material. The fabricated rectifier and antenna is connected with the help of RF cable known as rectenna. The fabrication of rectifier is shown in Fig. 10

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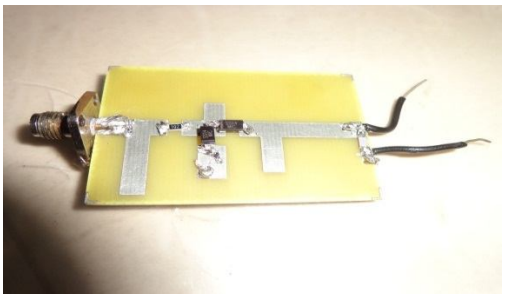

Fig. 10. Photograph of fabricated rectifier

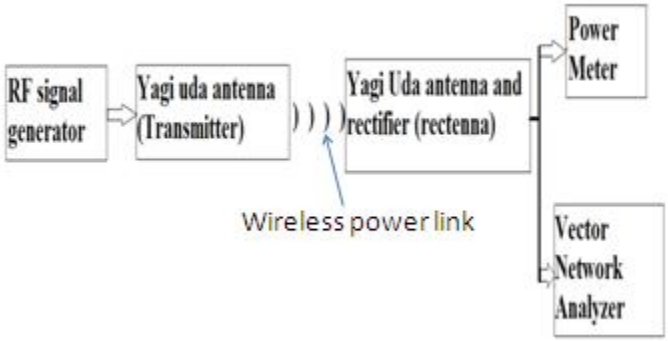

Fig. 11. Experimental setup for measurement

A. Measurement of Wireless Power Transfer

A Yagi Uda Antenna is placed in the transmitter side withgain of 8.14dBi at resonant frequency of 2.45GHz connected to the Agilent RF signal generator. In the receiver side a yagi uda antenna along with rectifier(rectenna) is placed. The input power fed to the yagi uda antenna is 10dBm. The distanceobtained between thistwo antenna is improved upto 2m. The power is transferred wirelessly between transmitter and receiver. The received power is measured using power meter. Measure the Dc output voltage of the rectenna and then estimated the total power transmission efficiency. The parameters such as VSWR, reflection coefficient, impedance is measured using Vector network analyzer(VNA). The measurement block diagram is shown in Fig. 11.The value obtained by 50Ω impedance matching of fabricated antenna is shown in Fig. 12.

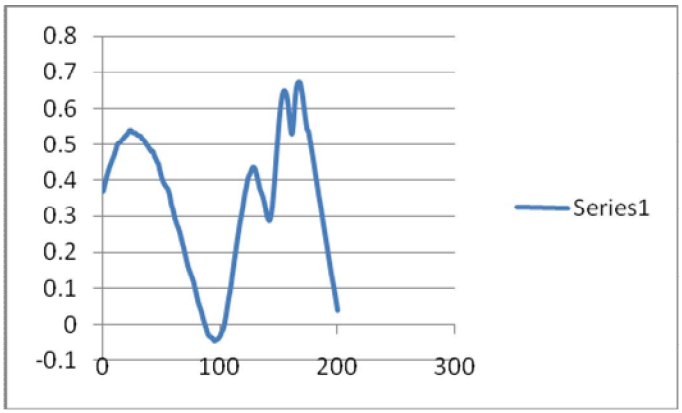

Fig. 12. Measurement of impedance mismatching using VNA

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The reflection coefficient of fabricated antenna achieved an return loss with three different frequencies is shown in Fig. 13. VSWR of fabricated antenna is shown in Fig. 14.

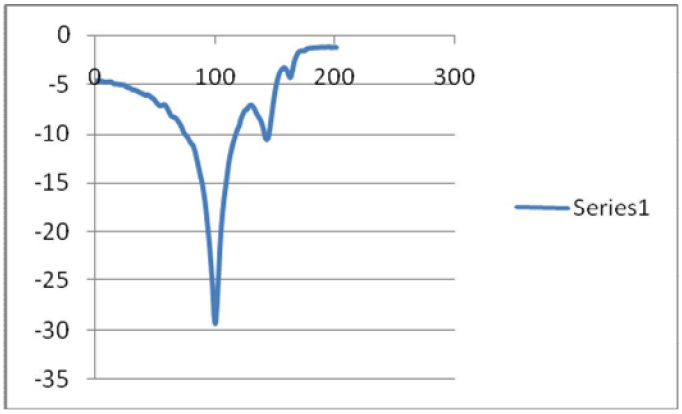

Fig. 13. Measurement of Reflection coefficient Vs return loss

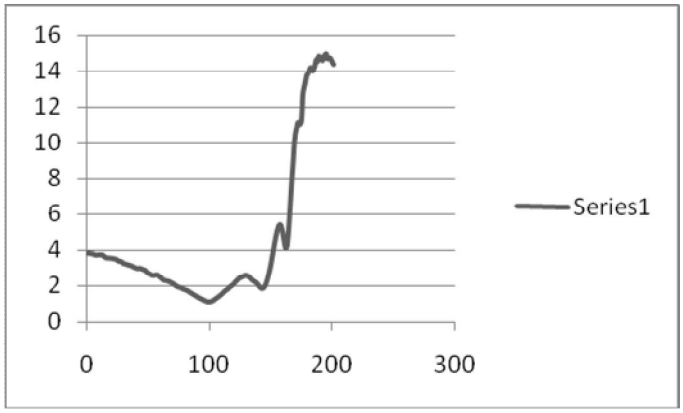

Fig. 14. Measurement of VSWR Vs Frequency

V. CONCLUSION

This paper presents a insitable rectenna for far field wireless power transfer at 2.45GHz. We first designed an yagi uda antenna and simulated using high frequency structure simulator with improved efficiency and gain. Secondly we designed rectifier for high RF to DC conversion HSMS2860 schottky diode.The value achieved between simulation and measurement is more or less similar. Further, frequency range can extend upto 20GHz, distance can be improved more and size of the rectenna can be miniaturize and it can be used for real time application.The simulated parameters is tabulated in Table given below

S.No.	Parameters	Microstrip Yagi Uda antenna
1.	Gain	8.14dB
2.	Directivity	9.19dB
3.	Efficiency	78%

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RADIATIVE POWER BEAMING USING MILLIMITER WAVES

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ABSTRACT

The concept of millimeter wave radiative power transmission in the W-band is presented. The power beaming system consists of a powerful source generating radiation at 94GHz, which is fed to a transmitting dish antenna. The transmitting antenna directs the electromagnetic radiation and focuses it onto a receiving rectifying array. The rectifying antenna (rectenna) converts the received radiation into DC electrical power, which is supplied to a consumer.

Various high power radiation sources have been used in the characterization of the rectenna. Solid states devices produce relatively low power (less than 400mW), while Free-Electron laser and Gyrotrons can produce several tens of kilo-Watts and above. This enables measurements of the rectifying antenna efficiency in different operational regimes.

We will discuss the expected performances of each element, including selection of optimal radiation frequencies as well as possible variable focusing transmitting antenna configuration. The characteristics of a complete power beaming at millimeter waves will be reviewed and design aspects of rectifying antenna array will be demonstrated.

I. INTRODUCTION

Wireless power transfer (WPT) techniques are being used to transfer electrical energy without physical wire connection between the power supply and the consumer. Radiative power beaming systems utilize high power electromagnetic radiation source. The electromagnetic energy is fed to a transmitting antenna which focuses the radiation onto a rectifying antenna at the receiving site [1]. A schematic illustration of a WPT system is shown in Figure 1. When millimeter waves (MMW) are used, electromagnetic wave may be generated by powerful radiation sources like gyrotrons [2] and free-electron masers [3]. The transmitting antenna is usually a dish in a cassegrain configuration, of which the focal length can be changed [4]. The rectenna is an array of patches connected to rectifying diodes that convert the MMW to a DC voltage [5].

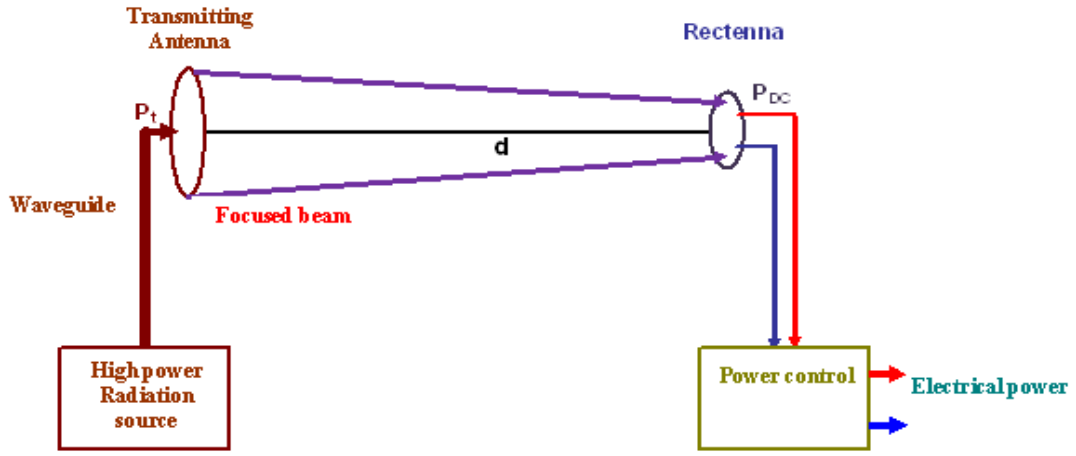


Figure 1: Schematic illustration of a radiative energy transmission system.

II. SOLID STATE TRANSMITTER

Several experiments have been carried out using different power sources at a frequency of 94 GHz. Continuous wave sources producing power of 0.4W and 2W where used. In these experiments, it was found that when the source strength is increased, the efficiency of the rectification increases. It should be noted that the power should not exceed a certain level which damages the rectifiers.

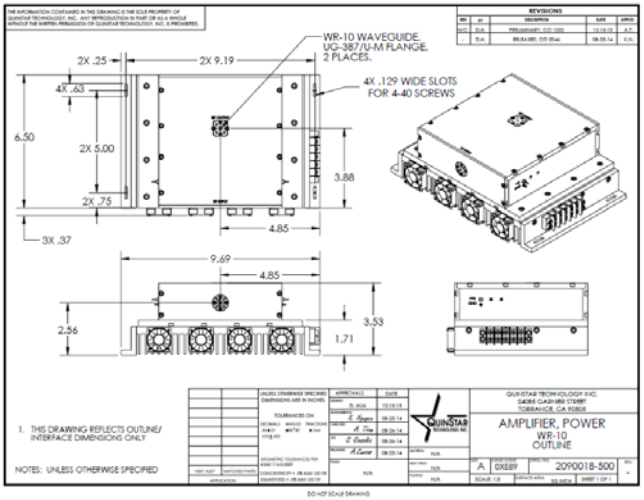
Increasing radiative power will improve the operational parameters of the system:

a. The DC voltage obtained after rectification will rise.

b. If storage capacitors are used, their charging time will be shortened

c. conversion efficiency of the rectenna will improve, while increasing the voltage obtained at the output.

In order to increase the power, a solid state amplifier was purchased from Quinstar. The amplifier model is QPN-94043314, operating at W-band and producing a continuous power of 2W. The Quinstar's amplifier is given in Figure 2.



Specification		
Frequency	92 to 96	GHz
Output Power (Psat)	33dBm min 32.5dBm typ	from 92 to 95 GHz @ 96 GHz
Gain	30 dB	min
VSWR	2 : 1	
Flatness	+/-2	dB
In/Out Waveguide Interface	WR-10	W/G
DC Supply	5V @ 10A typ	
Outline	2100005-500	Includes Heatsink w/ Fans

Figure 2: Quinstar solid stated power amplifier.

The experimental setup including the amplifier is shown in Figure 3. Several power beaming experiments have been conducted with the amplifier, showing improvement in the conversion efficiency.

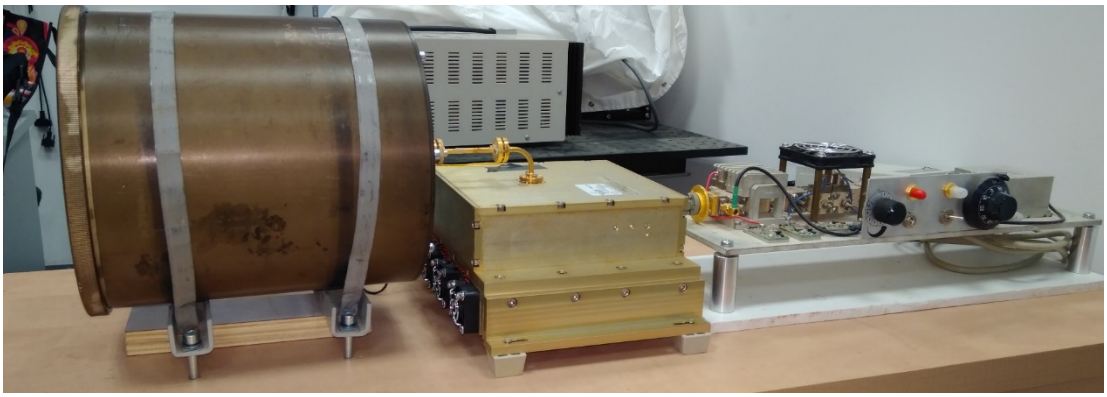


Figure 3: Power beaming solid-state transmitter.

III. THE RECTIFYING ANTENNA

The rectifying antenna consists of a 2x2 patches array connected to a single Schottky diode, as shown in Figure 4. Such diodes are used as detectors based on demonstrating high voltage responsivity, up to 16005 mV/mW at 87.8 GHz [6]. As reported in [5] these diodes can also be integrated with an mm-wave rectenna and operated in large-signal rectifying mode.

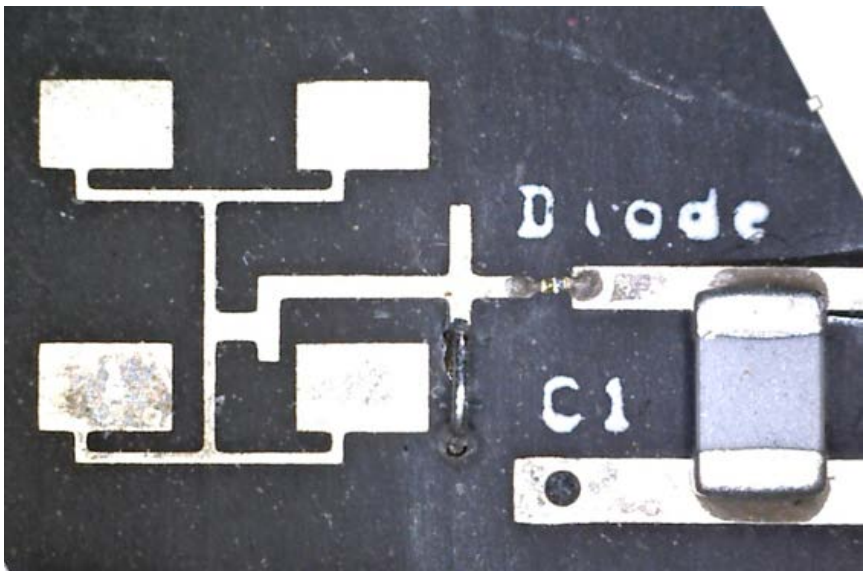


Figure 4: The fundamental 2x2 rectenna array.

In order to increase the received power, a design of 64 elements rectenna was made. A sub-array consisting of 4x8=32 patches was assembled using the 2x2 fundamental arrays, as depicted in Figure 5. Simulation of the rectifier and optimization of its performance has been done using ADS solver. The equivalent circuit used in simulation and typical results of ADS simulation are shown in Figure 6.

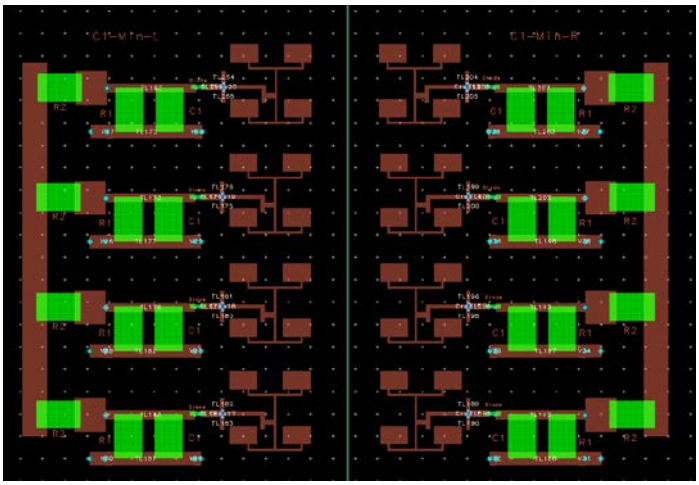
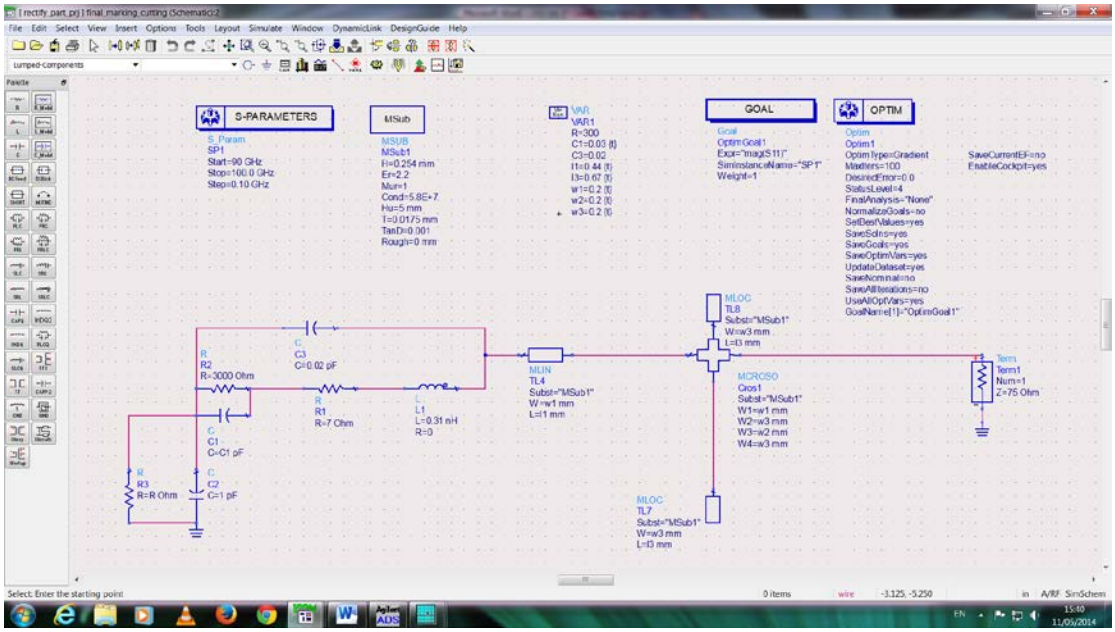
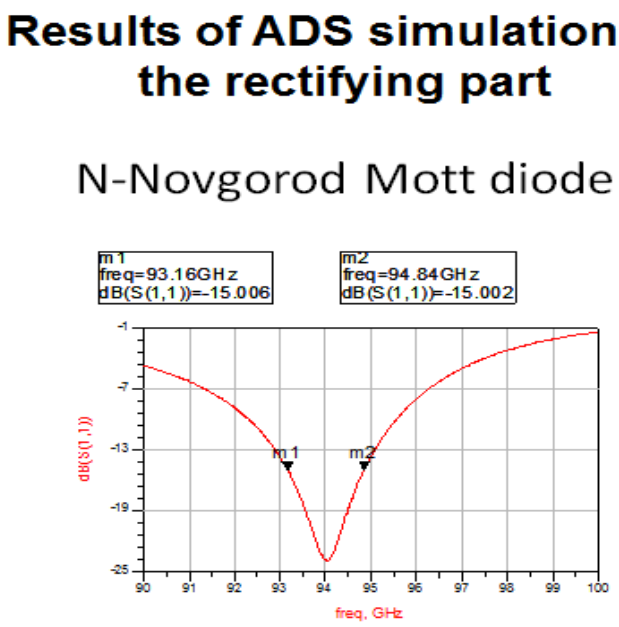


Figure 5: The sub-array rectenna consisting of 4x8 =32 patches.



(a)



(b)

Figure 6: The ADS simulation; a) The equivalent circuit, b) Reflectivity S11.

In the fundamental 2x2 array, the 4 patches are connected together to a single rectifying diode. The rectified waveform is filtered by a capacitor and fed to a resistor serving as a load. After optimization of the matching circuit to each of the diodes, the selected circuits were collected together to create a 64 patches rectenna as shown in Figure 7.

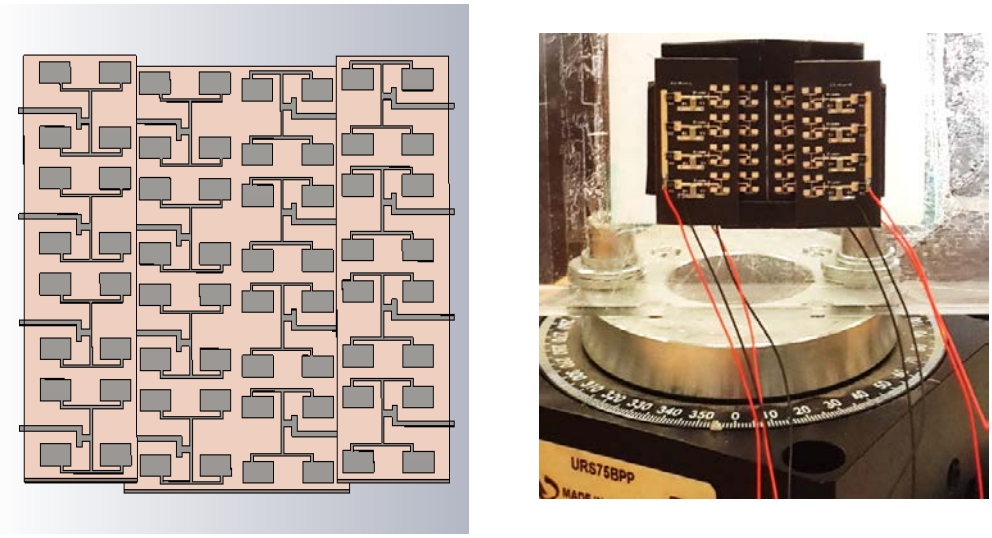


Figure 7: The layout of 64 elements rectenna.

IV. EXPERIMENTAL SETUP

Figure 8 presents the setup of the radiative power beaming experiment. A continuous wave, solid-state W-band source capable of producing 0.4W output power is connected to a 200mm lens transmitting antenna. The 2x2 fundamental array rectenna is placed behind the focusing lens of diameter 6 cm in a position of the beam spot, having a minimum diameter of approximately 1cm. distance between transmitting antenna and the receiving one is 1m. The graph of Figure 9 shows the measured DC power as a function of the load resistances. The maximum output power is 2.7mW for a load of 200 Ohms. The latter corresponds to the 15% of RF-to-DC conversion efficiency. This estimation was done using the technique reported in [5, 8].

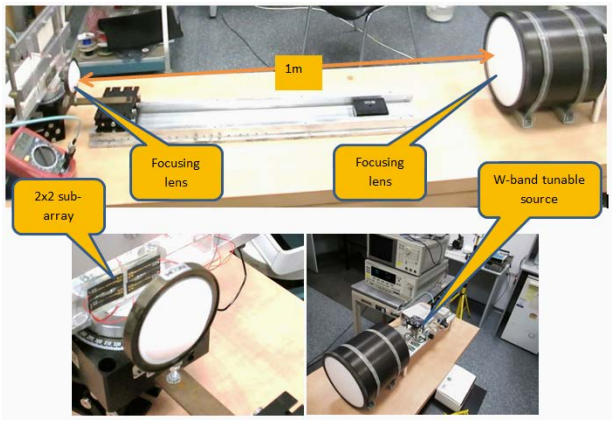


Figure 8: Experimental setup used for rectenna characterization in the W-band, using a 0.4W solid state tunable source

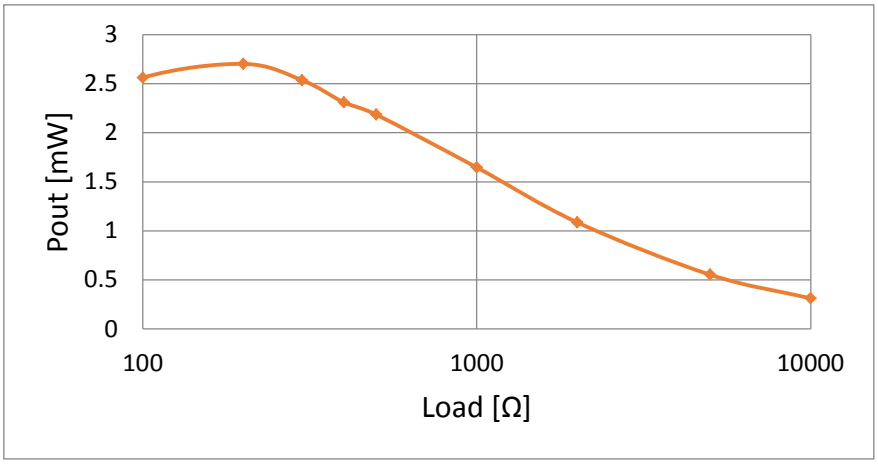


Figure 9: The measured DC power as a function of the load resistance.

V. SUMMARY

Millimeter wave radiative power beaming is demonstrated using solid-state power source and a rectifying antenna (rectenna). The rectenna is based on a fundamental 2x2 patches connected to a low-barrier Shottky diode. These modules are connected together to create a large rectifying array. Power beaming experiments at W-band, using continuous wave solid-state source demonstrate RF-to-DC conversion efficiency of 15-20%.

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Self-Powered Wireless Sensor

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Abstract

This paper develops a novel power harvesting system to harvest ambient RF energy to power a wireless sensor. Harvesting ambient RF energy is a very difficult task as the power levels are extremely weak. Simulation results show zero threshold MOSFETs are essential in the RF to DC conversion process. $0.5V_{DC}$ at the output of the RF to DC conversion stage is the minimum voltage which must be achieved for the micro-power sensor circuitry to operate. The weakest power level the proposed system can successfully harvest is -37dBm. The measured available power from the FM band has been measured to fluctuate between -33 to -43dBm using a ribbon FM dipole antenna. Ambient RF energy would best be used in conjunction with other forms of harvested ambient energy to increase diversity and dependability. The potential economic and environmental benefits make such endeavors truly worthwhile.

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1

1. Introduction

The contamination of landfills by batteries is a worldwide problem. Nearly three billion dry-cell batteries are consumed each year in the United States alone [1]. At the end of their operating life most batteries end in landfills. This contributes to a significant amount of hazardous compounds and heavy metals in the environment. To reduce reliance on batteries, energy harvesting techniques will be of vital importance in the near future. Energy harvesting is defined as the conversion of ambient energy into usable electrical energy [2-16].

The Citizen Eco-Drive is a famous example of energy harvesting. The watch collects ambient light and stores it as electrical energy in cells. These cells never need replacement and last longer than the life of the user [10], thus eliminating batteries ending up in landfills. There are many sources of ambient energy that can be harvested to perform work, such as solar power, wind power, energy from physical vibrations [12-15], thermal [11, 16], ocean tides and the RF spectrum. The primary focus of this investigation is to harvest energy directly from the RF spectrum. The RF spectrum covers the range from 3Hz to 300GHz [17]. The frequency band of interest is from the FM transmission band, it is located between 87.5 and 108 MHz of the RF spectrum [18]. There are two main reasons for selecting FM, firstly because of the extensive coverage and the polarization of the electro-magnetic fields are circular, which will enable the energy harvesting antenna to be positioned in virtually any orientation and still be able to receive a signal. The energy collected will be used to power a wireless temperature and humidity sensor node. The wireless and battery-less capability of these devices will greatly increase their utility.

The ultimate aim is to be capable of harvesting energy from across the entire RF spectrum, to maximize the amount of power collected. Additional investigation has been conducted, to further understanding on the challenges of harvesting RF energy at higher frequency levels. The frequency investigated is the GSM 900 cellular network. Historically, the attempts at harvesting energy from the cellular frequency band have only been of limited success as one is either restricted to transmitting their own power [19-20] or limited by the distance from the actual transmission tower [21].

This paper is structured as follows. Section 2 introduces the circuitry of the proposed RF power harvesting system. Section 3 provides an overview of the energy management system. Section 4 presents the energy harvesting performance of the prototype and also provides discussions on the results and the implications. Section 5 discusses future work is required. This paper is concluded in Section 6.

2. Energy Harvesting Circuitry

The proposed energy harvesting topology is shown in Figure 1.

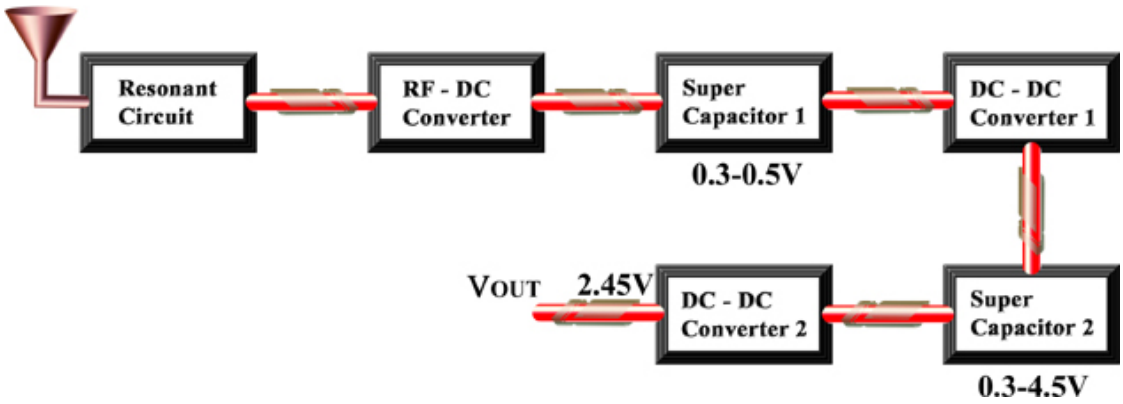

Figure 1*:* RF energy harvesting sequence

An impedance matching network has not been implemented between the energy harvesting antenna and the resonant circuit. This is because the energy harvesting circuit is a non-linear load, the input impedance changes with changes in voltage. Since the antenna and the input is not matched, reflection will always occur. The tuned series resonant circuit boosts the voltage amplitude of the incoming RF signal. The RF-DC converter rectifies the zero mean RF signal into DC and stores the charge on Super Capacitor 1. Super Capacitor 1 is used as a temporary storage capacitor, once the voltage reaches 0.5V, DC – DC Converter 1 transfers the energy to Super Capacitor 2 where the energy is stored at a higher voltage. This enables much more energy to be stored. DC-DC converter 2 is used to supply a constant 2.45V to the digital electronic components of the sensor node. By using a second DC-DC converter the voltage across Super Capacitor 2 can drop down to 0.3V before the sensor node will stop functioning.

2.1. Resonant Circuit

To increase the output voltage amplitude, the inductance is made as large as possible when compared with the capacitance value as shown in Figure 2. Surface mount components are used to reduce parasitic capacitance. The actual components are selected by their frequency response performance. The capacitors used are Class I ceramic capacitors. Their capacitance can be assumed to be frequency independent up to 1GHz [22]. The inductors used are high frequency multilayer chip inductors. The inductors are made from advanced ceramics with low resistance silver as the conductor [23]. The physical layout on the PCB is also important. To prevent the PCB from appearing as a transmission line, the components are placed as close together as possible. To reduce mutual inductance between inductors, the inductors are placed at 90 degrees to each other. For harvesting at FM transmission frequencies the resonant frequency is set at 100MHz.

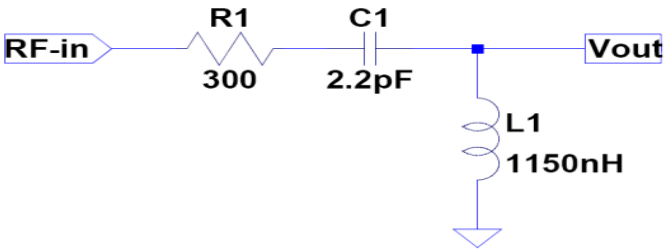

Figure 2*:* Series tuned resonant circuit

2.2. RF-DC Converter

The RF-DC converter is realized by cascading voltage doublers into a large voltage multiplier. In total 25 stages have been implemented. The Zero threshold MOSFETs are operating in the cut-off region, therefore the weak inversion current is utilized to rectify the RF signal as shown in Figure 3. Zero threshold MOSFETs reduce the voltage drop across each stage, therefore it increases the number of stages which can be cascaded. This is effectively Self-Synchronous Rectification (SSR).

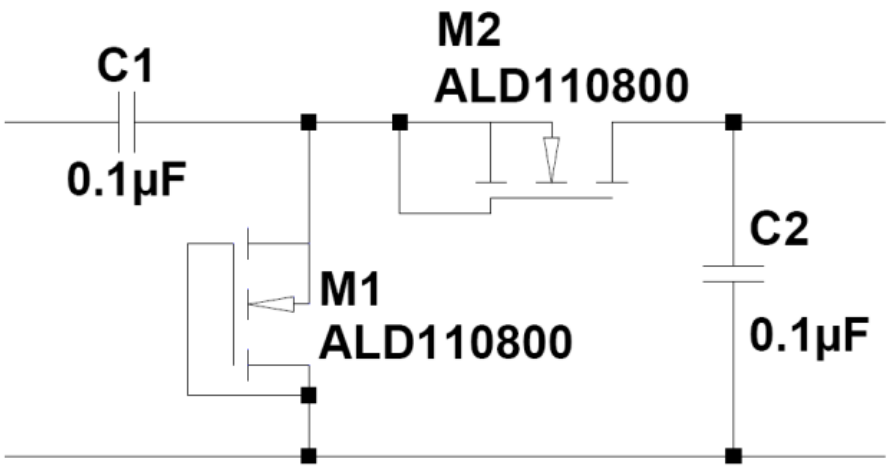

Figure 3*:* One voltage doubling stage

The capacitors used in the RF-DC converter are NP0 Class I ceramic capacitors. These are selected because of their excellent frequency response and the low leakage characteristics of the dielectric material. These are both very important qualities as the voltage multiplier is designed to work at frequencies in the MHz range. Also in order to improve efficiency any leakage current must be minimized.

2.3. Super Capacitor

Super capacitors are selected over conventional energy storage devices such as lithium-ion or lithium-ion polymer cells, because super capacitors have a close to infinite charging and discharging life cycles. There are two types of super capacitors used in the energy harvesting stage. The first super capacitor is used as a temporary energy storage device. Therefore an ordinary 1.5 Farad super capacitor is used to lower the cost of implementation. The second super capacitor is used to store the collected energy at a higher voltage to maximize the amount of energy which can be stored as shown by Equation 1.

$$E = \frac{1}{2} CV^2 \qquad (1)$$

The second super capacitor is a 1 Farad Bestcap, it has an ultra-low ESR, together with an extremely low leakage current [9]. A low leakage current is critical because harvesting ambient RF energy is very difficult therefore any charge collected is very valuable.

2.4. DC-DC Converter

The micro-power energy harvesting circuitry design contains two DC-DC converters. Existing available integrated chip converters were chosen over designing a converter from scratch because of several key reasons. The first reason is because of the very high efficiencies which can be achieved by using existing integrated converters. Efficiencies up to 90% can be easily attained. The second reason is a DC-DC converter which is able to accept a very large input voltage range is required. The input voltage range can vary between 0.3 to 4.5V. It is very difficult to manually implement a boost converter capable of operating with an input voltage of 0.3V. The final reason is integrated converters can automatically switch between buck and boost conversion thus simplifying development. The DC-DC converter selected for this application is the TPS61200 from Texas Instruments. It has a minimum startup voltage of 0.5V and can operate down to 0.3V. Both DC-DC converters are the same to simplify the design and manufacturing process.

3. Energy Management System

In Figure 4, the arrows show the direction of data flow. Some data lines are bi-directional. In Appendix B the schematic for the entire energy management system and the digital circuitry can be located.

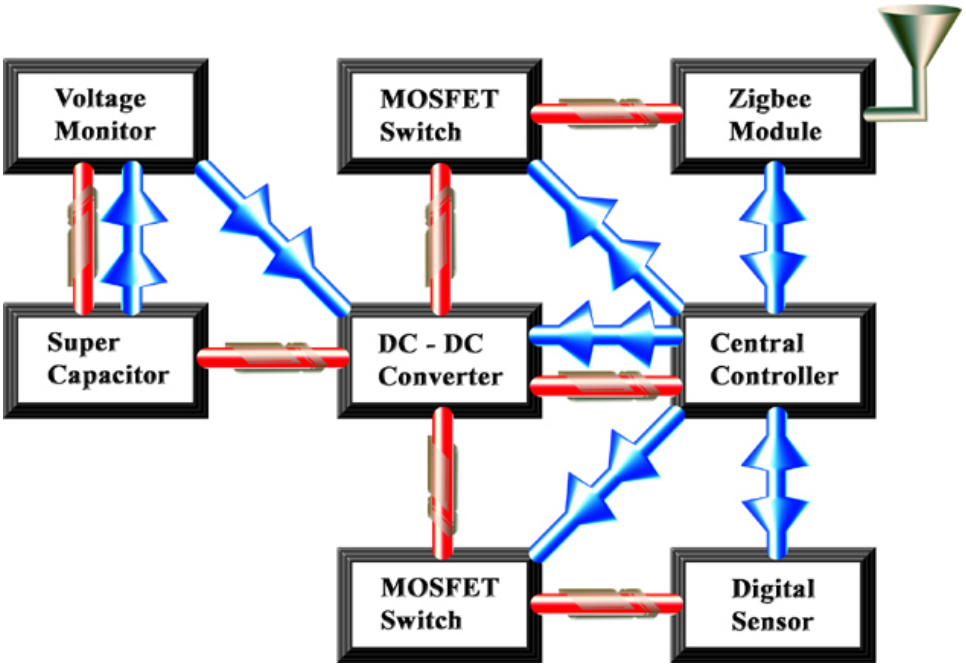

Figure 4: Sensor power and data flow

3.1. Voltage Monitor

The heart of the energy management system is the voltage monitor. It is implemented using an ATmega165P microcontroller. The voltage monitor operates at a clock frequency of 32.768 kHz and is powered directly from the super capacitor as shown in Figure 7. The voltage at which the voltage monitor starts to operate is 1.8V. The voltage monitor consumes 0.6 uA of current while in power-save mode. While in power-save mode, the real time clock is still operating. Once every week the voltage monitor will wake up from the power-save mode and check the voltage across the super capacitor. The regularity at which the voltage monitor checks the voltage stored on the super capacitor can be programmed to suit the particular application. Once sufficient voltage has been harvested, the voltage monitor enables the DC-DC converter which enables the central controller to operate. The voltage is checked through the ADC of the microcontroller. The code for operating the ADC is located in Appendix C. The ADC requires a voltage reference. The internal voltage

reference has been utilized to save on external components and can be turned on and off to conserve power.

3.2. Central Controller energy management

The central controller is also implemented using an ATmega165P microcontroller. It was selected because of its extremely low power consumption of a mere 10 uA in the active mode. Once the central controller powers up, the first priority is to gain control of the DC-DC converter. This is because the voltage monitor will return back to power-save mode to conserve energy. Without control of the DC-DC converter the central controller will have no power source. To gain control of the DC-DC converter the central controller must set the I/O pin which is connected to the enable pin of the DC-DC converter to high. The central controller is capable of monitoring the voltage across the super capacitor. This enables the controller to know how much time is left for the tasks to be completed. Once the data has been transmitted back to the data recorder, the central controller will disable the DC-DC converter.

3.3. Power MOSFET switch

Although both the zigbee module and the digital sensor have sleep modes, they still consume current. To eliminate this unnecessary leakage current, MOSFET switches have been added to allow the central controller full control of when each device will receive power. This enables precise timing to be achieved thus minimizing the amount of power used. The MOSFET used are the IRLML2502PbF from International Rectifier. These are selected because it has a maximum threshold voltage of 1.2V, this is essential as the operating voltage within the sensor node is 2.45V.

Another advantage with the IRLML2502PbF MOSFET it has a very low ON resistance of 0.045 ohms. Both the zigbee module and the digital sensor node have a MOSFET switch connected.

4. Results and Discussions

The power levels were measured using a spectrum analyzer. Two antennas have been tested at FM transmission levels. The first antenna is a length of insulated copper wire which has been cut to half a wavelength, to resemble a monopole antenna. A SMA connector is used to connect this antenna to the spectrum analyzer. The amount of power measured at the output of the antenna is -50dBm. The second antenna is a standard ribbon FM dipole antenna. The power measured at the output varied between -33dBm and -43dBm. This antenna is shown in Figure 5.

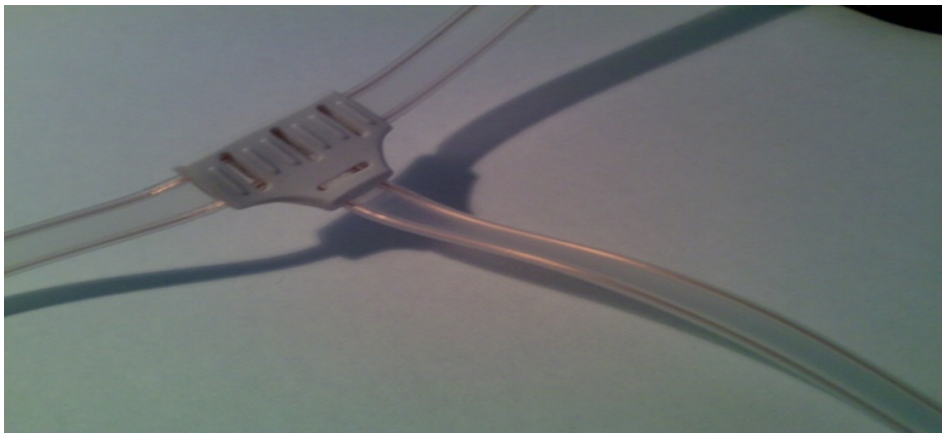

Figure 5*:* Ribbon FM dipole antenna

The procedure used to test the RF energy harvesting performance of a given circuit is as follows. The circuit input is connected to the output of a precision signal generator. The signal generators are capable of varying the amount of power in dBm and also the signal frequency. $0.5V_{DC}$ at the output of the RF to DC conversion stage is the minimum voltage which must be achieved for the micro-power sensor circuitry to operate. $0.5V_{DC}$ is therefore the standard by which we compare the two different rectifying schemes.

A total of 20 cascaded voltage doubling stages have been built to test the performance of Schottky diodes in rectifying small signals. The smallest power that could be rectified to 0.5V was -18dBm. It is also worthy to note after 7 to 8 doubling stages have been cascaded, the voltage gain of adding additional stages diminishes. To further improve the energy harvesting circuit, a series resonant circuit was added to boost the amplitude of the incoming RF signal. The hardware implementation was successful. A 9 dB voltage gain has been achieved. However this is at the cost of compromising the RF bandwidth from which we are harvesting. The top and bottom layers of the prototype are shown in Figures 6 and 7, respectively.

The energy required for each device is based on how long each device is operating and the amount of current each device draws. Power calculations for a typical wireless sensor application are tabulated in Table 1. It is shown that the energy required to complete one data transmission cycle is 0.32 Joules. The result has been confirmed by measuring the voltage difference before and after data transmission.

Device	**V (V)**	**I (mA)**	**T (s)**	**E (J)**
ATmega165P	1.8	0.01	10	0.00018
ATmega165P	1.8	0.01	8	0.00014
Sensor	3.3	0.55	5	0.00908
Zigbee	3.3	35	2.7	0.31
Total				**0.32**

Table 1*:* Power Calculations for a wireless sensor

Assuming there is a constant -37dBm power flow entering the energy harvesting antenna. It will take 11.7 days for the RF harvester to collect 0.32 Joules of energy. This does not take into account the reflection which may occur between the antenna and the input of the RF harvester,

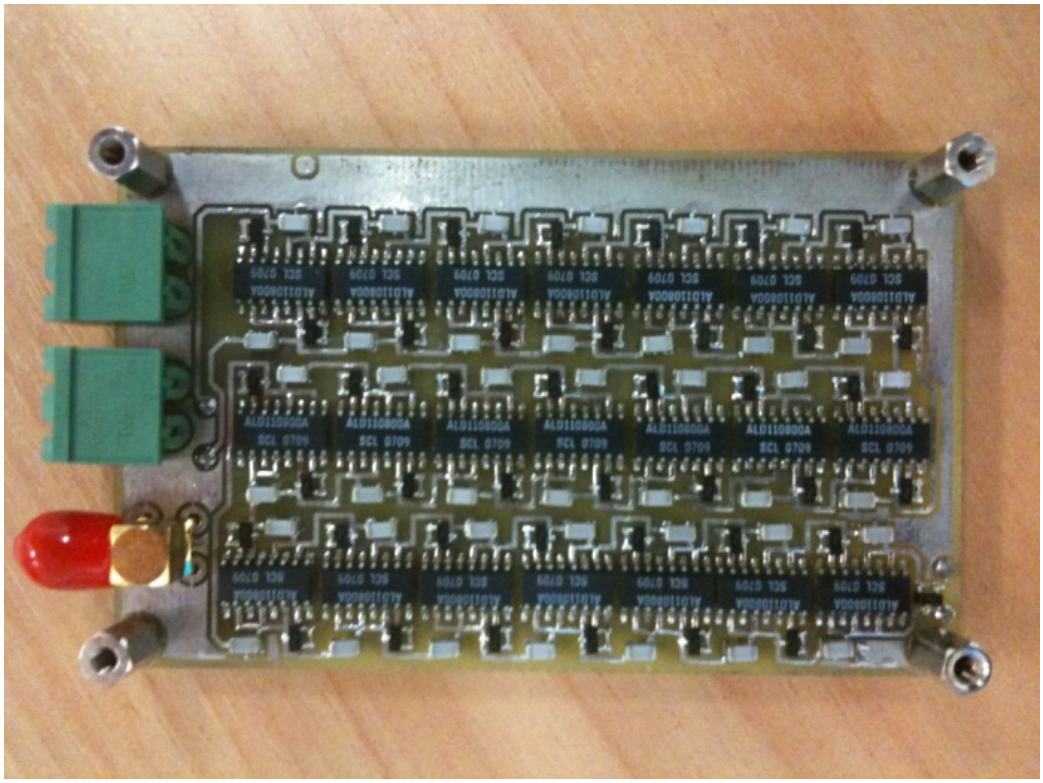

Figure 6: Power harvesting system (top layer)

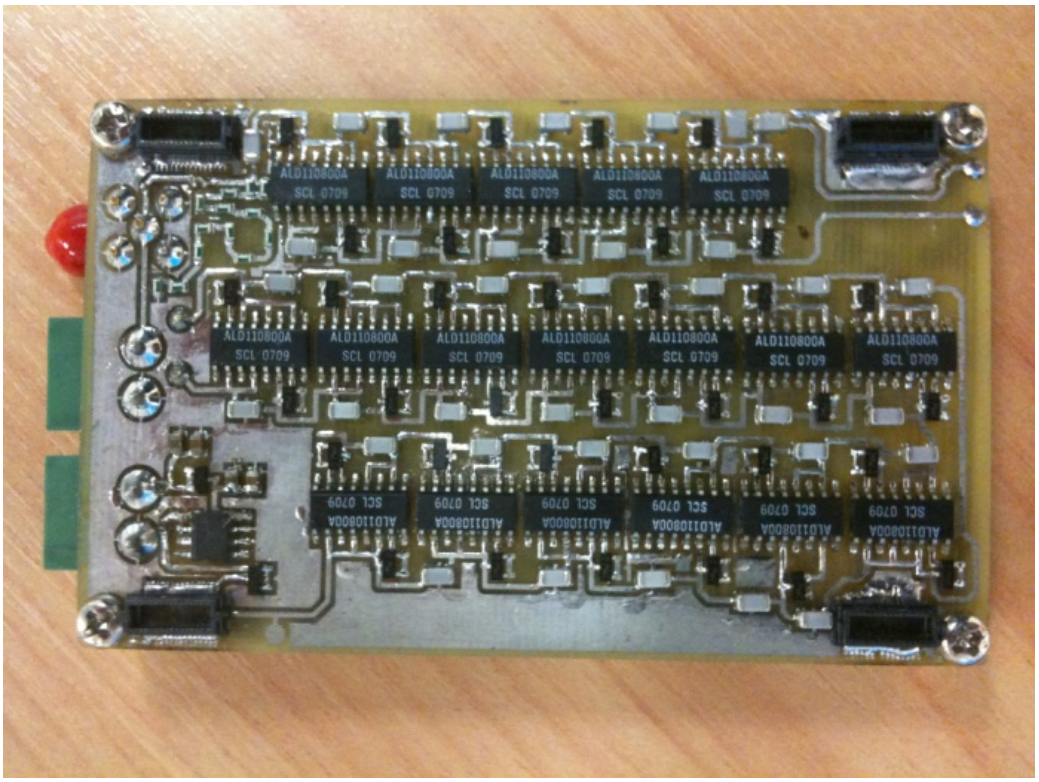

Figure 7: Power harvesting system (bottom layer)

The realistic time the RF energy harvesting circuit will take to collect enough energy for one data transmission will be closer to 20 to 30 days. This is taking into account the reflection coefficient and the fluctuations in power levels. There are still many applications where such a device will be very useful. One such application is in silver-culture. Here circumferential sensors are used to measure the growth of trees. Since the trees grow very slowly the time between measurements is usually a month or more. By powering these sensors from an ambient source, it eliminates the need to ever change the batteries. For applications requiring faster charge rates, it is possible to combine energy harvested from various different ambient sources to increase the available power and diversity.

5. Future Developments

To be truly successful at harvesting ambient RF energy, the entire energy harvesting circuitry must be integrated on a single IC. This is to reduce leakage losses, increases the overall efficiency of the system and help energy harvesting from the RF spectrum to be economically viable. For this project the objective was to harvest only from the FM transmission band, however the power collected is very limited and the time taken to harvest the energy is very long. Therefore the next step is to increase the bandwidth from which the RF energy harvester can collect energy. This is no trivial task, because in radio systems it is easy design a circuit to work at a certain frequency but for the circuit to work at multiple frequencies may require different harvesting techniques. It will be necessary to design a broadband antenna. Another area which needs improvement is the power sensitivity. It is highly probable that by making the entire circuitry physically smaller the system will be capable of harvesting even lower power levels. A possible future development is to integrate RF energy harvesting with other forms of ambient energy such heat and physical vibrations.

6. Conclusions

Harvesting ambient RF energy is very difficult as the power levels are very weak. Schottky diode based voltage multipliers are not suitable for rectifying such weak signals. A tuned series resonant circuit enables the energy harvesting circuitry to harvest at lower power levels but the bandwidth is compromised. The real hardware implementation supports the simulation results showing that SSR using n-MOSFETs is essential in the RF to DC conversion process. The lowest power level which can be boosted and rectified is -37dBm at 100MHz. At 900MHz the lowest power level the energy harvesting circuit can harvest increased to -25dBm therefore the harvesting performance at higher

frequencies is not as good. Since the measured available power from a FM ribbon dipole antenna fluctuated between -33 and -43dBm, harvesting energy from the FM spectrum it feasible.

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A Novel RF Energy Harvesting Module Integrated on a Single Substrate

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ABSTRACT

This paper presents the RF energy harvesting module (RECTENNA). The working range of this module includes multiple bands i.e. GSM, ISM, WLAN, and UWB band. To enhance the capturing RF power capability an array arrangement of coplanar monopole antenna has been proposed. Wilkinson power combiner has also been implemented to combine the powers of this antenna array. The RF-DC converter circuit having seven stages has also been integrated with this structure. This module produces the DC voltage of 1.8 V with respect to + 40 dB RF input. It is the unique module because it has no need of port connectors. The impedance matching of antenna and converter has been fulfilled by incorporating the passive component at the combiner's branch. The value of this passive component is kept equal to the existing value of impedance at input port of converter circuit.

Key words: *Rectenna, coplanar monopole antenna, Wilkinson power combiner, antenna array, impedance matching, RF-DC converter circuit.*

I. INTRODUCTION

The primary focus of the Rectenna circuit is to design a suitable RF receiving antenna. For this purpose, an antenna should be designed to resonate on single, dual and multiband frequencies according to the need of the applications. The main aim of proposed work is to design a RF energy harvesting module for a wide range of frequencies which covers most of functional bands, e.g. GSM, Radio, ISM and UWB. It is well known that the RF spectrum widely employed for 900 MHz – 2 MHz (television and radio applications, GSM), 2.1GHz - 2.6 GHz (ISM band for various applications), and 3.1GHz -10.6 GHz (ultra wideband for satellite applications) [1]. Narrow band systems such as WLAN (3.1GHz - 4.4GHz), HIPERLAN (5.1GHz - 5.3GHz), C-BAND (4.4GHz - 5GHz) are also the useful bands. Keeping this in mind a compact coplanar monopole antenna has been designed and its 2 X 2 arrangement was reported [2]. For preparing the array arrangements of the antenna the literature [3-5] were profoundly studied.

This antenna works for the wide range of frequencies of the above specified bands. As reported in the structure [2] the Wilkinson power combiners were preferred because it accomplishes two essential tasks of the structure. Firstly it combines the powers of two or four antennas serially and increases the input power of RF-DC converter circuit of energy harvesting module. Secondly it considered as the element of afford, for the impedance matching of antenna port and converter circuit. The maximum power transfer concept has been chosen for this impedance matching.

The multi band operation of the RF energy harvesting module developed on a single substrate is the innovation of this paper.

II. ANTENNA DESIGN

In RF energy harvesting module antenna is the major functionary part. It receives the power from the electromagnetic spectrum. The received RF power is converted into the DC output voltage from the module. So the antenna should be capable enough to receive the maximum energy/power from the spectrum. As the working range of our design is preferred from 900 MHz -10 GHz, the modification in the monopole antenna structure for multi frequencies resonance has investigated. For increasing the area of receiving as well as gain of the structure the 2 x 2 array of the coplanar monopole antenna has been designed and presented [2]. In this paper [2] only the simulated results were presented. The simulated antenna structure was as shown in figure 1[2]. The power of the individual structure has been combined by the Wilkinson power combiner. The reason for selecting this combiner is that we may externally adjoin the surface mount components at the combining branch of it for impedance matching purpose.

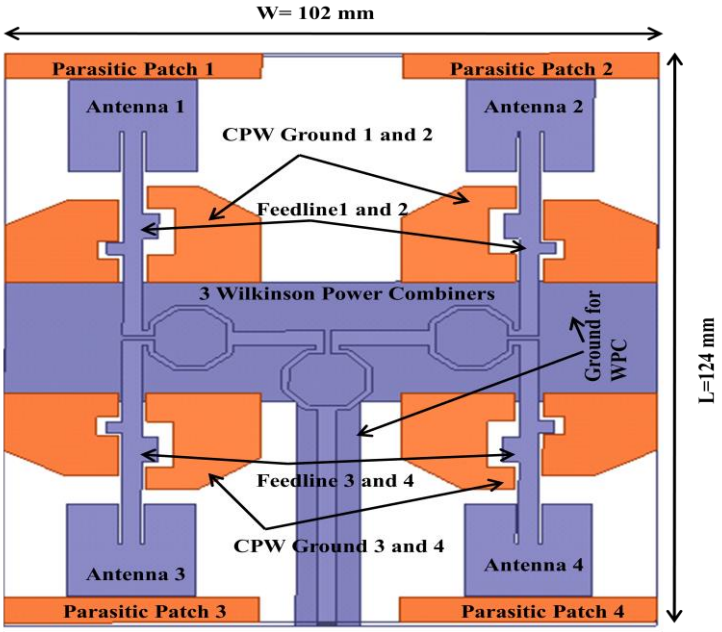

Figure 1. Simulated structure of 2 x 2 array arrangement of coplanar monopole antenna [2].

The simulated structure from licensed software HFSS version16 has been exposed resonant on the multi frequencies of different allocated bands as 900 MHz and 1.75 GHz (GSM band) 2.7 GHz (ISM), 3.5 GHz, 4 GHz and 5.2 GHz (WLAN), 6.5 GHz, 7.4 GHz, 8.5 GHz and 10 GHz (UWB).

Further in this paper all measured results have been reported for the validation of proposed antenna structure. For the measurement purpose proposed structure has been fabricated as shown in figure 2. The measured result of return loss from VNA (Keysight Technologies) for fabricated structure is shown in figure 3. The resonant frequencies of the fabricated structure has

exposed on 900 MHz (GSM), 1.29 GHz (ISM), 4.1 GHz and 5.6 GHz (WLAN), 6.8 GHz and 9 GHz (UWB), that is the almost same as simulated result of the structure reported earlier [2].

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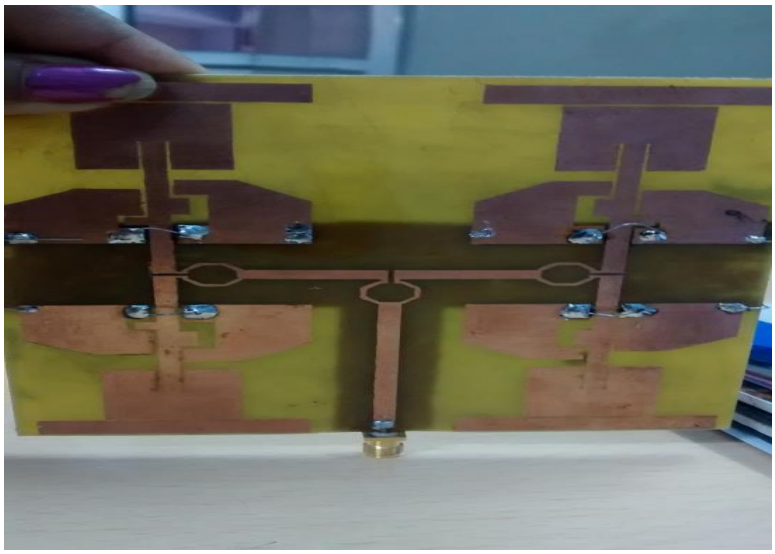

Figure 2. Fabricated structure of the 2 x 2 array of coplanar monopole antenna.

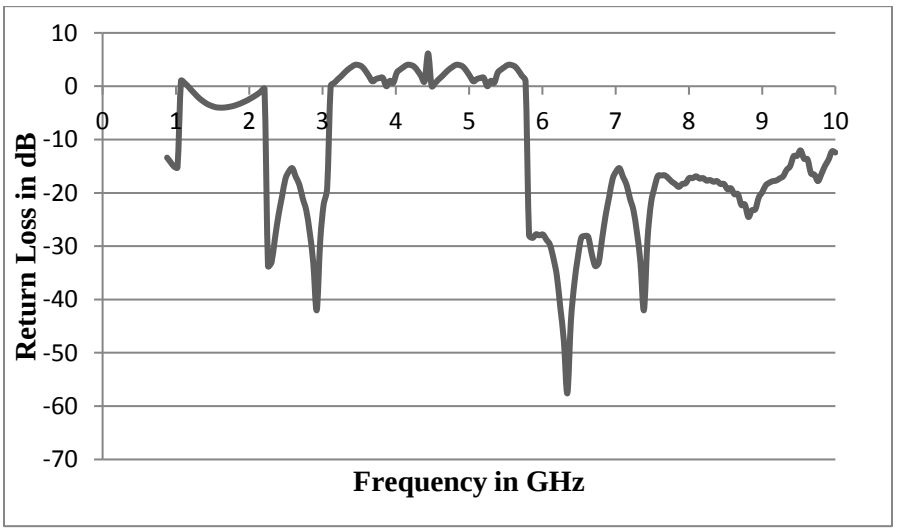

Figure 3. Measured return loss (S_{11}) of the 2 x 2 array of coplanar monopole antenna.

The measured H- field patterns for these frequencies are shown in figure 4(a) and E-field patterns are shown in figure 4 (b). These patterns show that the antenna supposed to be worked as omnidirectional.

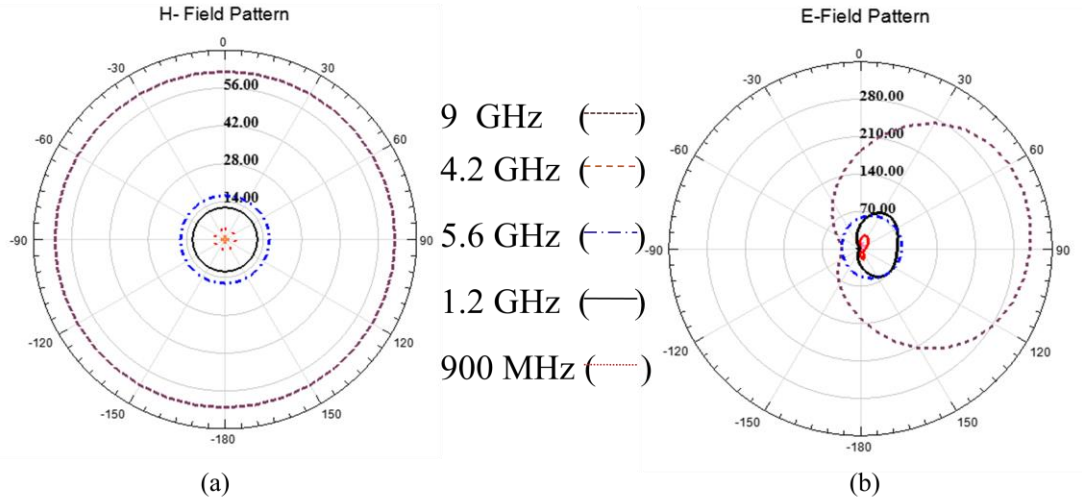

Figure 4. (a) H- field pattern (b) E- field Pattern

III. RF-DC CONVERTER

The RF-DC converter circuit is also the important part of energy harvesting module. The output of harvester circuit is the DC voltage. The input of this module is the magnitude of AC voltage at the antenna port. This AC voltage is proportional to the captured power on the antenna from the RF spectrum. In this paper for energy conversion purpose a voltage doubler circuit has been used. The number of stages of the converter circuit may be increased for escalating the DC output voltage value. The relation of number of stages (n) and output voltage ($V_{o/p\ volt}$) of converter circuit is considered as

$$V_{o/p\ volt} = \frac{nV_{o/p\ volt}}{nR_0 + R_L} R_L$$

In this paper conversion design contains seven numbers of stages of Villard voltage doubler circuit. The main function of this conversion circuit is to convert the RF harvested energy from the ambient into the direct current (DC) voltage. This seven stage Schottky diode voltage doubler circuit is first designed, modelled and simulated on ANSYS RF circuit designer and then fabricated and tested. After that, this circuit has integrated with the same substrate of antenna. The converter circuit in this design uses zero bias Schottky diode (SMS7621-001-SOT23) from Skyworks. The simulated design of seven stage RF-DC converter circuit has been presented for this purpose [6]. The simulated DC output voltage at the first stage and seventh stage of RF-DC converter circuit from the RF circuit simulator was as 48.2 mV and 289.6 mV respectively. The fabricated circuit of seven stage RF-DC converter and its analysis has already been presented [6]. So in this paper we implemented that designed RF-DC circuit (having seven stages) [6] with the same single substrate of designed antenna array.

IV. RF ENERGY HARVESTING MODULE

This part of paper presented a RF energy harvesting module on a single substrate working on multi bands of the frequency range 900 MHz to 10 GHz. The DC output from the multiband harvesting circuit has been presented it produces 3.4 V DC output [8]. The motive of this part is to design whole module on a single substrate for a wide range of operation and should be able to produce the DC output voltage of the value more than 1.5 V from the ambient RF spectrum. This DC voltage may be boosted by adding the booster circuit. In addition to the structure on a single substrate of RF module, it is also proposed that any value of impedance measured at the converter circuit input impedance may be matched to antenna port by using SMD (Surface mount device) resistor components. This is possible only due to the use of Wilkinson power combiner implemented at the antenna structure. No need of matching of connectors impedances. So it may be called connector-less module. This is the uniqueness of the structure.

The novel idea of impedance matching is that; for any wilkinson power combiner a given number of output ports N, an input impedance R_S and output load impedance R_L, and the characteristic impedance of each quarter-wave section, $Z_{\lambda/4}$, is given by

$$Z_{\lambda/4} = \sqrt{N\, R_L R_S}$$

Additionally

$$R = R_L$$

the resistor (R) that connect between the common junction and the output ports are simply equal to the load impedances (R_L) at each output. Now considering the input impedance of RF-DC converter circuit as the value of load resistance (R_L) of the output of the Wilkinson power combiner. This impedance has been measured by vector network analyzer. The SMD resistor of this measured value (R) has been mounted in between the combining leg of Wilkinson power combiner. In this way we can easily match the impedance of antenna and converter circuit. This satisfies the condition of maximum power transfer.

The fabricated RF energy module on a single substrate with impedance matching arrangement is shown in figure 5. The seven stage RF-DC converter has been implemented by the combination of seven Schottky diodes and seven capacitors. The load resistance and source resistance has also been added into the circuit. The overall size of the module is 124 x 102 mm^2.

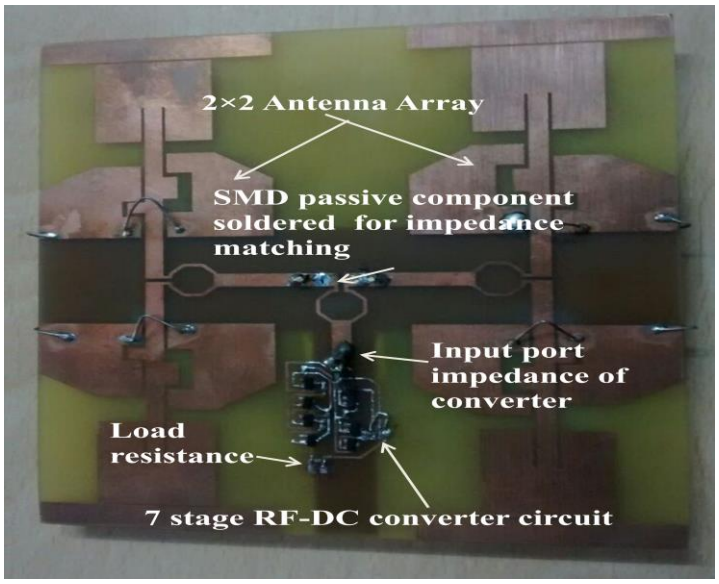

Figure 5. Fabricated RF energy harvesting module integrated on a single substrate.

The measurement with module in RF lab has been shown in figure 6. The measurement has taken at ICRS (International center for Radio Science) lab, Jodhpur. Rajasthan, INDIA. The RF source is of 40 GHz. First of all with respect to the source on the resonant frequencies of the module; power received at the antenna port has been measured on the power-meter.

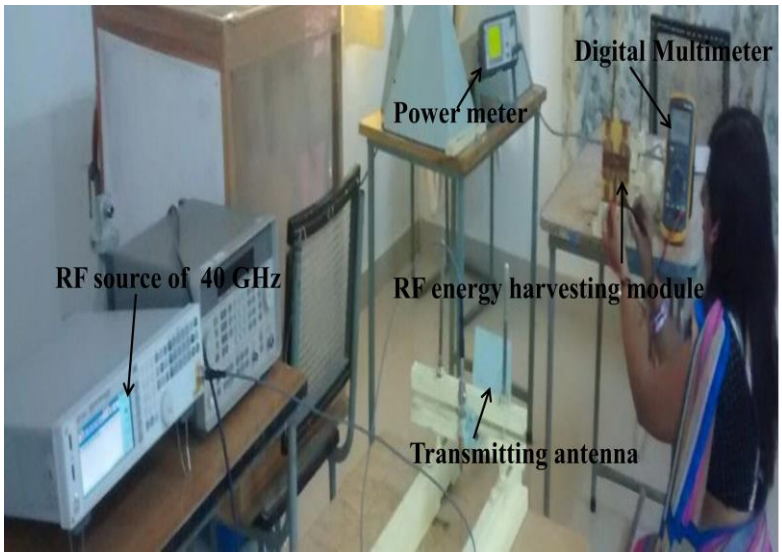

Figure 6. Measurement of output DC voltage from the module with respect to RF input voltage from RF source.

The DC output voltage at the module output with respect to RF input power varying from - 40 dB m to + 40 dB m has been plotted for five resonant frequencies (900 MHz, 1.2 GHz, 4.1 GHz,

5.6 GHz and 9 GHz). The plot of the data is as shown in figure 7. The maximum DC output is 1.823 V.

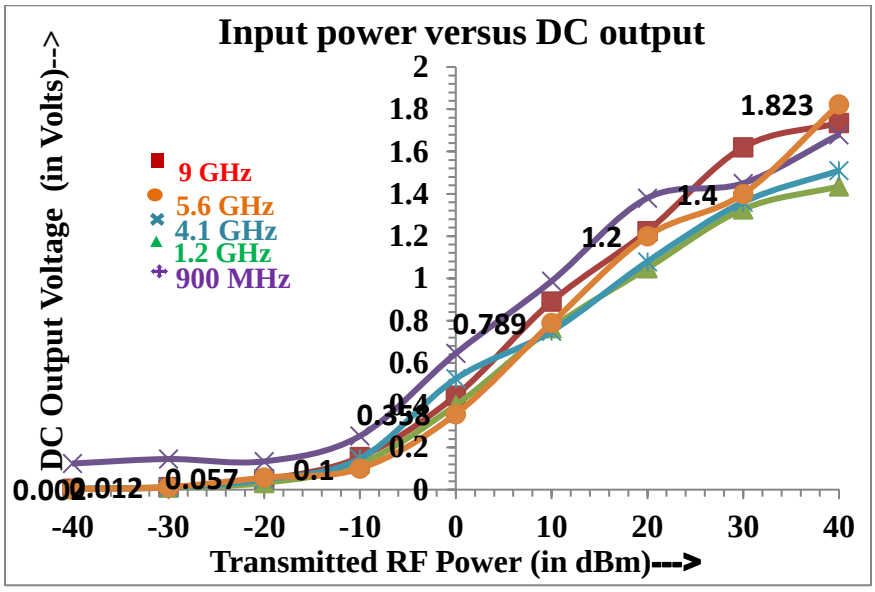

Figure 7. The transmitted RF input power versus DC output at the module graph.

The conversion efficiency (in %)of the module is calculated by the formula

$$\eta = \frac{V_{DC\ output}}{V_{input}} \times 100$$

The $V_{DC\ output}$ is the output measured by the digital multimeter and V_{input} is the magnitude of the input voltage at RF power source. The efficiency for the 9 GHz frequency is shown in the figure 8. As shown in figure the efficiency increases as the RF power increases.

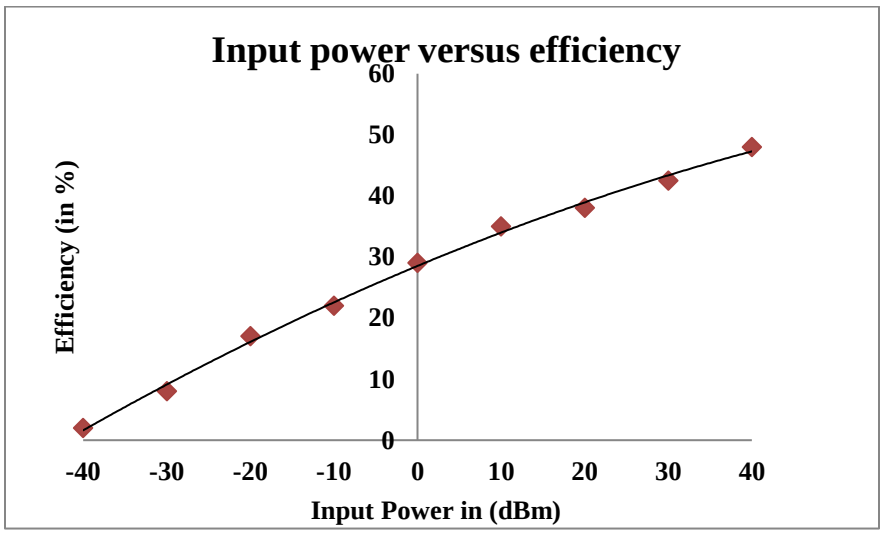

Figure 8. Input RF power versus efficiency graph.

The efficiency increases by the change in load resistance. For the input RF power +10 dBm the change in efficiency with respect to the change in load resistance is shown in figure 9. The maximum DC output voltage (1.823 V) for +40 dBm has been noted on the load resistance 22 k-ohms.

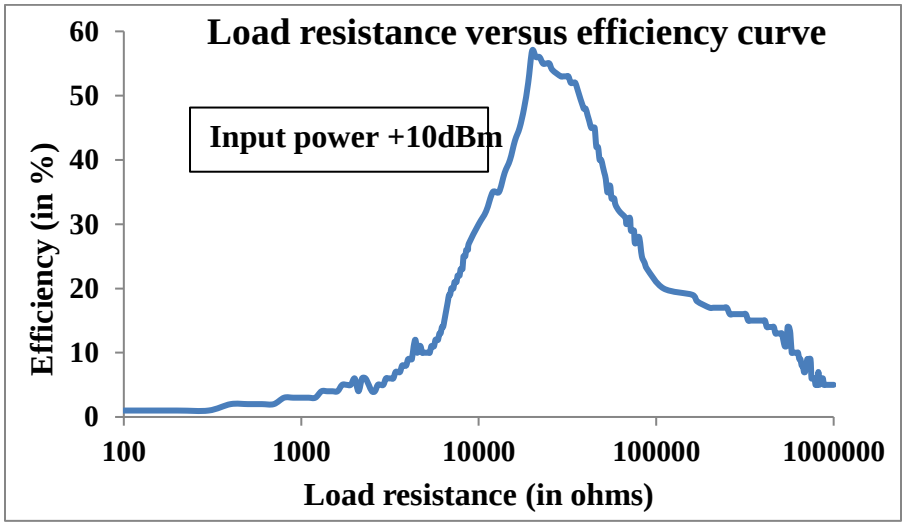

Figure 9. The load resistance versus efficiency curve.

V. CONCLUSION

A novel RF energy harvesting module has been presented in this paper. The antenna for this module has the structure of 2× 2 array of the coplanar monopole antenna. The novelty which is proposed lies in the fact that a seven stage RF-DC converter circuit has been integrated on the same substrate of antenna. This makes the module connector-less. Another uniqueness of this module is that the impedance between the antenna and the converter circuit is matched by using the maximum power transfer concept. The Wilkinson power combiners have been specifically selected for this purpose as they combine the powers of antenna arrays. For impedance matching between the antenna output port and RF-DC converter port, the existing input impedance at the input port of the converter circuit is measured. And the component of that measured value has been mounted at the Wilkinson combining branch. This arrangement reported the matching between the antenna and circuit ports. The DC output of the module has been measured. The maximum DC output is 1.823 V with respect to +40 dB m input RF power, with the load resistance 22 k ohm for the frequency 9 GHz. This voltage may be boosted by integrating a booster circuit. The voltage may be increased by using the super capacitor at the output of module.

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Practical Rectennas
Far-Field RF Power Harvesting and Transport

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Practical Rectennas
Far-Field RF Power Harvesting and Transport

PROEFSCHRIFT

ter verkrijging van de graad van doctor aan de
Technische Universiteit Eindhoven, op gezag van de
rector magnificus, prof.dr.ir. C.J. van Duijn, voor een
commissie aangewezen door het College voor
Promoties in het openbaar te verdedigen
op donderdag 5 juni 2014 om 16.00 uur

door

Shady Keyrouz

geboren te Bsharri, Libanon

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"It is only with the heart that one can see rightly,
what is essential is invisible to the eye."

Antoine de Saint-Exupry (1900-1944)

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Summary

Practical Rectennas: Far-Field RF Power Harvesting and Transport

Energy harvesting from renewable sources such as the sun and wind, as well as from other sources like heat and vibration have been the focus of research recently. However, an increasing number of applications, especially inside buildings where light/solar energy is not always sufficient, cannot profit from these sources. Wireless Power Transmission (WPT) overcomes the lack of different power sources, such as light, heat and vibration and is an attractive powering method for battery-less sensor nodes, and active Radio-Frequency Identification (RFID) tags.
In this thesis, different rectenna topologies are investigated and state-of-the-art rectenna systems are discussed. The basic far-field energy harvester (rectenna) consists of an antenna, which captures a part of a transmitted RF signal, that is connected to a rectifier which converts the AC captured signal into a DC power signal. For a maximum power transfer, a matching network between the antenna and the rectifier is needed. In order to decrease the size of the system and improve the power conversion efficiency, the matching network can be removed by directly matching the antenna to the rectifier. A power management circuit is needed after the rectifier to store the harvested DC power and regulate the output voltage. This thesis is divided into four main parts:

- The rectifier is investigated in the first part. Since the rectifier is a nonlinear device and its impedance changes as a function of frequency and power, an accurate analysis model of the rectifier's impedance as a function of both is presented. The model is verified using Harmonic Balance simulations and measurements. In addition, the optimization of the rectifier's parameters for maximizing the output voltage is presented.

- The second part of the thesis presents a far-field power transport system employ-

ing rectennas using matching networks. A novel, modified, 50Ω Yagi-Uda antenna is presented. The main advantage of the presented antenna is that it can operate as a broad-band, dual band or a multi-band antenna. Three prototypes are fabricated, and the simulation results are validated by measurements. A rectenna system operating at 868 MHz is experimentally validated. Results are reported, demonstrating, at an input power level of -10 dBm, an RF to DC power conversion efficiency (PCE) of 49.7%. The 50Ω rectenna system exhibits an increased PCE of at least 5 % at -10 dBm and at least 10 % at -20 dBm, compared to reported state-of-the-art results.

- The focus of the third part of the thesis is on reducing the antenna size and at the same time keeping a high radiation efficiency. A novel (modified) rectangular loop antenna is presented. The benefits of the suggested novel antenna topology are the compact size and the ability to tune its input impedance while keeping a constant radiation efficiency. Using the suggested antenna, at an input power level of -10 dBm, the power conversion efficiency reaches 54%. The novel compact rectenna system is smaller in size and exhibits an increased PCE of at least 10 % at -10 dBm compared to reported state-of-the-art results.

- The fourth part of the thesis introduces a method to harvest RF power using Frequency Selective Surfaces (FSS). The novel FSS design is capable to harvest power from arbitrarily polarized incident waves. The design does not involve a matching network which results in a simple, polarization-independent power harvester. A simple equivalent circuit model to predict the transmission and reflection characteristics of the gridded-square-loop FSS is employed. Two prototypes have been fabricated and an RF to DC conversion efficiency of 25 % for the first prototype and 15.9 % for the second one are measured at $3.2\mu\,\mathrm{W/cm^2}$ available power density level.

Recommendations, future work, and means to enhance the power conversion efficiency as well as the system performance are addressed in the conclusion chapter.

Chapter one

Introduction

1.1 Introduction

In this chapter we will give the rationale for the analysis and development of *rect*ifying ant*ennas* or *rectennas*. Rectennas are used for converting RF energy, radiated by a distant (intentional or unintentional) source, into usable DC energy. Rectennas are used for transferring far-field, radiative RF energy into DC energy and in that respect differ from inductive or non-radiative, resonant, RF energy transferring systems. The differences between the two ways of RF-to-DC power conversion will be given and potential applications for far-field RF energy transfer will be discussed. These applications will be linked to wireless sensors. The outline of the thesis will be discussed, highlighting the new and original contributions. The chapter will be concluded by an overview of frequency bands used throughout the thesis and a single-number Figure of Merit used to compare complex numbers resulting from various simulations and measurements.

1.1.1 Trends and Needs

The use of wireless devices is rapidly expanding. The number of mobile phone subscriptions worldwide, for example, has expanded more than twelve times between 1997 and 2007, see Fig. 1.1 [1].

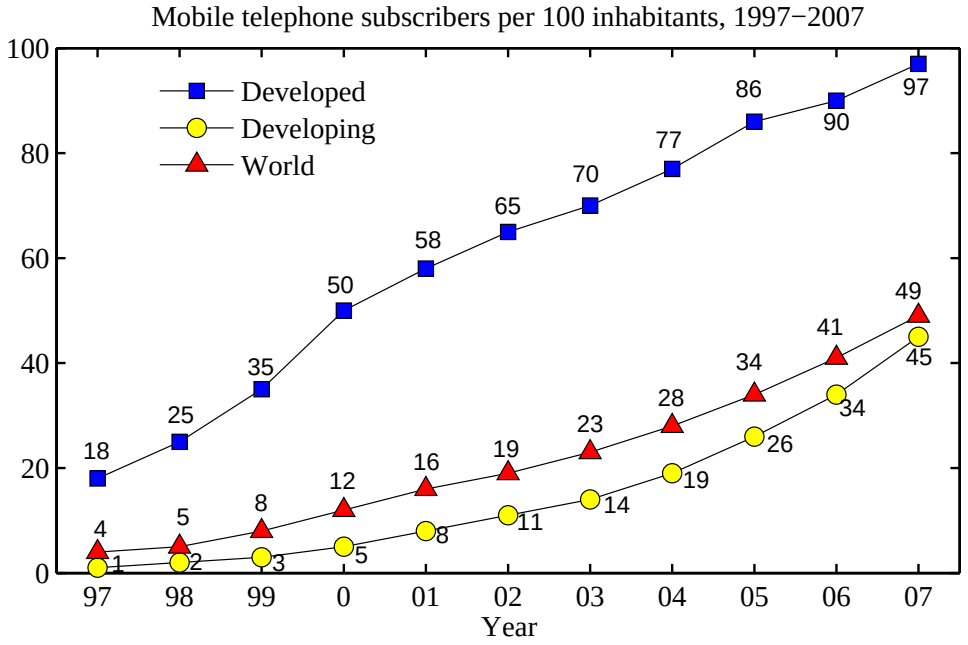


Figure 1.1. Mobile phone subscribers per 100 inhabitants, 1997-2007 [1]. Courtesy ITU .

Both Cisco [2] and Ericsson [3] expect 50 billion wireless devices to be connected to the internet in 2020, serving a population of 7.6 billion [2]. For wireless devices in general, the world Wireless Research Forum (WWRF) estimates 7 trillion wireless devices serving 7 billion people [4]. The total market for wireless sensor devices totaled $532 million in 2010, $790 million in 2011 and is expected to increase to $4.7 billion in 2016 [5].

Next to this expected growth in the use of wireless devices, also a need is developing for an increased wireless sensor autonomy and small form factor, see Fig. 1.2 [6]. The need for long lifetime and small form factors of future wireless sensor nodes do not match up well with currently available battery technology [7]. Alternative ways of powering may be found in energy harvesting where energy is taken from ambient sources like vibration, thermal differences, light and radiofrequency (RF) signals.

One of the main potential applications we see for employing far-field RF energy harvesting and transfer is in the so-called Smart Building Initiative (SBI) concept [8]. Before discussing SBI in detail, we will first discuss the differences between inductive resonant-

coupling and radiative far-field power transfer and the differences between transfer and harvesting.

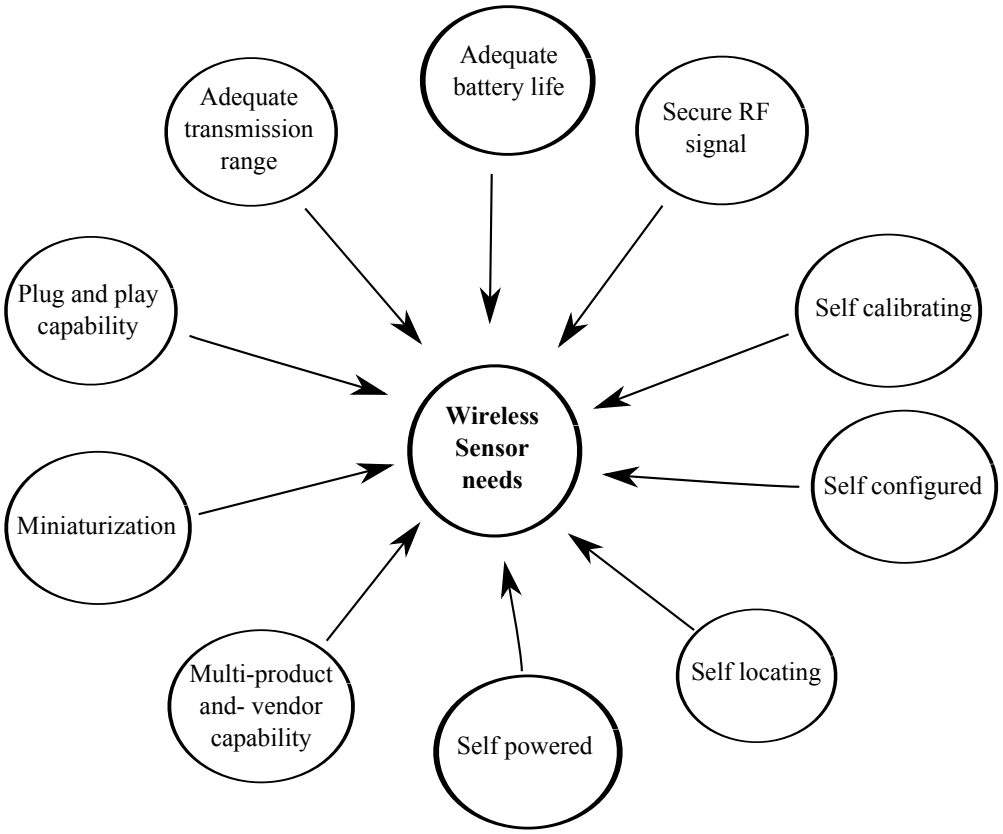


Figure 1.2. Needs for wireless sensors.

1.2 Non-Radiative and Radiative RF Power Transfer

When employing RF signals for the transfer of energy, we make a distinction between non-radiative and radiative RF energy transfer. The former is based on (inductive) coupling, the latter uses the transmission and reception of radio waves. When the transmitting source is unintentional, we call the transfer of energy *harvesting*. For a radiative system this might take the form of converting e.g. the energy in radio broadcasting signals into DC energy.

1.2.1 Non-Radiative Transfer

Non-radiative RF energy transfer is mostly employed in inductive systems, obeying the Qi standard [9]. Two coils are brought in each others vicinity for transferring RF energy, see Fig. 1.3. The two coils in close proximity form an electrical transformer. By

applying capacitors, a resonant coupling system is created [10], increasing the power transfer efficiency.

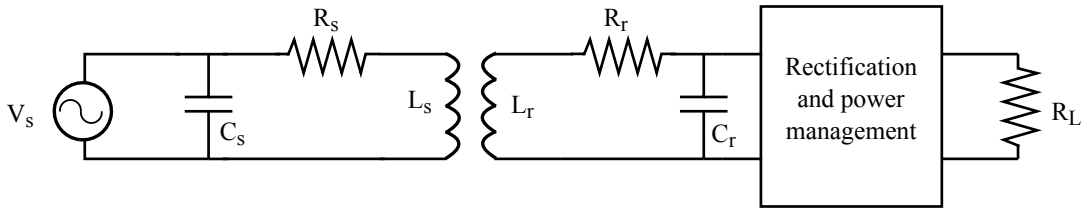


Figure 1.3. Resonant coupling (non-radiative) RF energy transfer system.

In the circuit shown in Fig. 1.3, a RF source is connected to a coil L_s, that is resonated with a capacitor C_s. A receiving coil L_r, resonated with a capacitor C_r is placed in close proximity and the received signal is rectified, eventually accumulated and used to power a source that is represented by load resistance R_L. The resistances R_s and R_r represent the losses in the transmitting and receiving coils, respectively. The benefit of this resonant coupling energy transfer, apart from being subject to a standard, lies in the high power transfer efficiency [11]. The drawbacks are in the limited transferring distance, the transfer efficiency falls off very steeply for distances larger than approximately one coil diameter, which also makes harvesting (taking energy from the ambient) impossible. For RF energy transfer over a distance, we need to employ radiative transfer.

1.2.2 Radiative Transfer

In a radiative transfer system, a RF source (unintentional or intentional) is connected to an antenna that emits radio waves. At a distance from this radio source, a receiving antenna intercepts a part of the radiated waves and the RF signal captured by the antenna is rectified, eventually accumulated and transferred to a load (wireless sensor), see Fig. 1.4. The dashed box in this figure is in fact a *rect*ifying ant*enna*, commonly known as a rectenna.

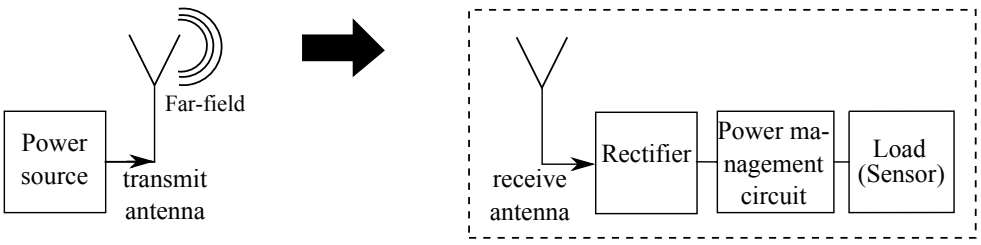


Figure 1.4. Radiative RF energy transfer system. The dashed box is commonly denoted rectenna.

Wireless radio-frequency (RF) energy transmission dates back to the experiments of Heinrich Hertz, in the 1880s, proving Maxwells theory of electromagnetics [12]. Initiated by an interest in short-range ($\leq$ 2 m [13]) radio-frequency identification (RFID) applications, far-field RF energy transfer has become a research topic at many European, American and Asian universities and research institutes [14–22]

1.3 Smart Building Initiative (SBI)

In a smart building, wireless sensors will measure, amongst others, the presence of persons, the local temperature and the lighting strength. Based on these readings that are transmitted to a central station, lighting and heating conditions will be adjusted (switched off or reduced if no persons are present). Thus energy consumption will be reduced. Artificial lighting, for instance, contributes approximately 30% of electricity consumption in a commercial building, and buildings collectively account for 40% of total world energy use [23]. SBI may also add to an increased well-being of elderly people when employed in an assisted living scenario, allowing the aged to live at home for a longer period [24], [25]. SBI relies on the use of a large number of wireless sensors, see also Fig. 1.5 [26].

Figure 1.5. Wireless sensors in a smart building.

Both aspects of SBI, a reduction in energy consumption and assisted living, are expected to generate a large economic impulse. The European home automation market revenues

were US$232.6 million in 2007 with an expected increase to US$446.6 million in 2013 [27]. In April 2011, the European smart homes and assisted living market was estimated to grow by 16.2% from US$1,544 million in 2010 to US$3,267 million in 2015 [28]. Recent estimates of revenues from home automation systems in Europe and North America project a growth at a compound annual rate of 42% from US$ 1.2 billion in 2012 to nearly US$ 9.4 billion in 2017 [29].

This needed widespread use of wireless sensors in offices and houses has, however, been severely slowed down due to the cost of cabling. Cabling an average office unit or a four bed house is estimated at €7,000–11,000 ($9,000–15,000) [30]. This is especially true for offices, that are, on average, reconfigured every five years [31]. Although using batteries may severely reduce these cabling costs, the maintenance issues associated with using primary batteries – i.e. locating batteries to be replaced and disposing of the old ones will not be tolerated by the market [32].

Then what remains for powering the many wireless sensors is energy harvesting. Well known sources for energy harvesting are [7]: vibration, ambient light, temperature gradients, and (ambient) radio waves. With the exception of intentional radio waves, the availability of energy harvesting sources of sufficient power cannot be guaranteed under all circumstances [12]. Therefore we will concentrate on the use of dedicated RF sources, distributed in a building for powering the wireless sensors. The investment in these sources and the costs of powering them will be more than compensated by the reduction in power consumption of the whole building by switching off lights and turning down heating.

1.4 Challenges

For powering the wireless sensors needed in SBI we will employ the license-free frequency bands for Industry, Science and Medicine (ISM). Although these frequency bands can be used without the need for a license, they are restricted in the maximum allowed Effective Isotropic Radiated Power (EIRP), that is the transmit power multiplied by the transmit antenna gain. For ISM frequencies between 865 MHz and 2.48 GHz the maximum allowed EIRP does not exceed 4W [12]. Due to the spherical spreading of the transmitted electromagnetic waves, the power received at the rectenna side will decrease inversely with the squared distance between transmitter and receiver so that, in general, we will have to deal with low received RF power levels. That means that all the subsystems of a rectenna, especially the antenna and the rectifier need to be designed and impedance matched very carefully. As already demonstrated in [33], the RF-to-DC power conversion efficiency of the rectenna may be maximized by directly, complex-conjugate impedance matching the antenna to the rectifier. Thus, we need to be able to determine the (complex) RF input impedance of a rectifier circuit and need

to be able to model and design antennas generating a desired complex input impedance. Furthermore, we need to be able to find the optimum input power level and frequency to operate the rectifier circuit. All these topics will be discussed in this thesis.

1.5 Organization of the Thesis and Original and Novel Contributions

The organization of the thesis is graphically depicted in Fig. 1.6.

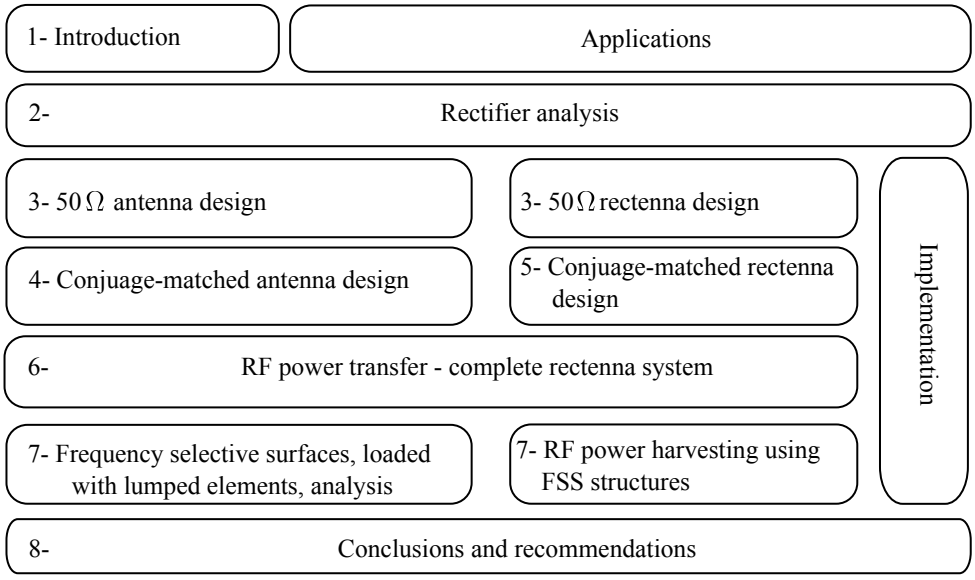


Figure 1.6. Outline of this thesis.

In Chapter 1 the problem of powering an increasing number of wireless devices is discussed as well as the possibilities of harvesting power from the ambient. The use of RF power transfer and harvesting is discussed in more detail and far-field, radiative RF power transfer is chosen as the preferred way to transfer power wirelessly over a substantial distance. The main application of this technology is seen in Smart Building Initiatives and assisted living. Due to restricted power densities it is concluded that the design of the subsystems of a rectifying antenna (rectenna) as well as the impedance matching between these subsystems must be carried out very carefully. Therefore it is important to develop accurate and easy to implement and use models of especially rectifying circuits and antennas. On top of that, a design strategy with respect to the choice of the operating frequency, RF input power and impedance matching needs to be developed. In Chapter 2, the analytical modeling of a high-frequency, rectifying Schottky diode is discussed. Based on the model, the influence of the bulk resistance and the

saturation current on the DC output voltage is analyzed. The main contributions and innovations of Chapter 2 are:

- The derivation of a Schottky diode selection criterion and the selection of commercially available Schottky diodes according to this criterion.

Chapters 3 to 5 deal with antenna design, where the designs are immediately applied in a co-design with a rectifier to create rectenna structures, see Fig. 1.6. Chapter 3 deals with an antenna structure that can be designed for dual, triple or broadband frequency use. Based on a reported broadband strip-dipole antenna, a Yagi-Uda like configuration is created around this active strip-dipole antenna to improve the antenna characteristics with respect to bandwidth and gain. The antenna is designed for a 50 Ω input impedance. The new configuration is used to design a rectenna, using a lumped-element network to match the antenna impedance to the complex input impedance of a voltage doubling rectifier.

The main contributions and innovations of Chapter 3 are:

- A wideband strip-dipole antenna is combined with parasitic elements in a Yagi-Uda like antenna configuration to create a dual, triple or broadband antenna having a moderate gain (3 to 6 dBi).
- The new antenna concept has been used to design a rectenna that demonstrates an increased Power Conversion Efficiency (PCE) of at least 5 % at -10 dBm and at least 10 % at -20 dBm with respect to the state of the art as published in the open literature.

Chapter 4 deals with antenna structures intended for a direct conjugate matching to the rectifier, thus omitting an impedance matching network that will lead to a higher power conversion efficiency (less loss) and a more compact design (fewer lumped elements). Started is in chapter 4 with the strip-dipole antenna analysis. The derived model is used in modeling a strip folded-dipole antenna. Evolving from this latter antenna, a modified loop antenna is introduced and full-wave analyzed. The main contributions and innovations of Chapter 4 are:

- The validity range (length and width) of the impedance equation for a strip dipole antenna has been extended by deriving new fitting equations.
- The accuracy of an existing strip-folded dipole antenna has been improved by using the improved strip-dipole antenna model.
- A new, electrically small antenna type has been invented that allows for tuning the complex input impedance and the radiation efficiency independently.

In Chapter 5 a design strategy for maximizing the sensitivity of a rectenna is outlined. As a consequence, an antenna should be designed for having a very low real part of the input impedance and a high absolute value of the imaginary part. The new antenna discussed in chapter 4 is used for this purpose and is co-designed with a lumped-element Schottky diode rectifier and with an integrated custom-made rectifier. Both realized rectennas demonstrate characteristics better than the state of the art. The main contributions and innovations of Chapter 5 are:

- The realization of a rectenna with dimensions smaller than the state of the art and demonstrating a RF-to-DC power conversion efficiency larger than state-of-the art.
- A rectenna based on a co-design of the same antenna and a custom made rectifier demonstrating the smallest dimensions paired to the highest sensitivity ever reported in the open literature.

In Chapter 6, a complete rectenna system is presented where a commercially available power management circuit, a 1000 μF storage capacitor and a dynamic load are used. The average DC output power and the system's RF to DC conversion efficiency are measured, and compared to those measured for a commercially available rectenna system operated under the same conditions. It is shown that the commercially available rectenna system fails to operate at low power levels (≤ -6dBm) while the presented system can operate at an RF input power level as low as -13 dBm. At an input power level of 0 dBm, the power conversion efficiency of the commercially available system does not exceed 19 %, while the power conversion efficiency of the presented system reaches 40 %, resulting in more than 20 % improved power conversion efficiency. The main contributions and innovations of Chapter 6 are:

- Full rectenna system prototyped and characterized.
- Improved performance compared to commercially available system.

In Chapter 7, the analysis of a gridded square loop Frequency Selective Surface (FSS) is extended with the inclusion of lumped elements to model the presence of Schottky diodes for harvesting RF power. The updated model that is validated through cross-simulations and waveguide simulator measurements predicts the resonance frequencies with the diodes included and thus makes it possible to design RF harvesting FSS structures. The main contributions and innovations of Chapter 7 are:

- The recognition of the harvesting structure being essentially a FSS.
- The inclusion of square loops in a grid to maximize the current density in that grid.

- The modeling of the RF power harvesting FSS and the use of this model to optimize a large area RF power harvesting structure.

Finally, Chapter 8 presents a summary of the main results, conclusions and recommendations for future research.

1.6 Figure of Merit

In the comparison of different simulation and measurement results for a complex impedance throughout this thesis, we want to end up with a single-number Figure of Merit. For that purpose the power wave reflection coefficient is an excellent tool, relating to a practical use [33].

1.6.1 Power Wave Reflection Coefficient

When a source with internal impedance Z_R is connected to a load impedance Z_L, see Fig. 1.7, the standard (voltage) reflection coefficient Γ is given by

$$\Gamma = \frac{Z_L - Z_R}{Z_L + Z_R}. \tag{1.1}$$

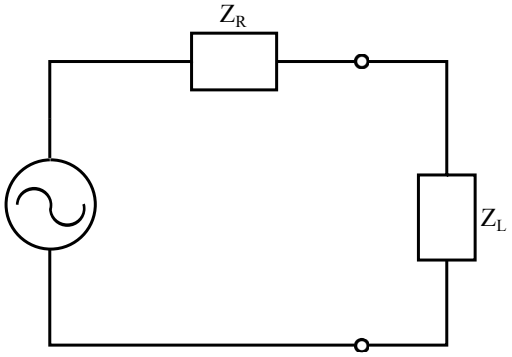


Figure 1.7. A source with internal impedance Z_R is connected to a load impedance Z_L.

This reflection coefficient is a direct measure for the reflected power only when Z_R is real-valued [34]. When Z_R is complex and conjugate impedance matching is applied to maximize the power transfer from generator to load, the standard reflection coefficient in Eq. 1.1 will be nonzero. Therefore, the standard definition of reflection coefficient, based on physical waves propagating along a transmission line, does not in general

represent the reflection of power. Power waves, mathematical constructs from the circuit world, have been introduced to correctly describe the reflection of power in systems with complex reference impedances. The reflection coefficient based on power waves Γ_p is given by

$$\Gamma_p = \frac{Z_L - Z_R^*}{Z_L + Z_R}, \tag{1.2}$$

where Z_R^* denotes the complex conjugate of Z_R. The reflection coefficient according to Eq. 1.2 works for both real-valued and complex impedances.

1.6.2 Application of the Power Wave Reflection Coefficient as a Figure of Merit

We will now apply this power wave reflection coefficient as a Figure of Merit (FoM) to describe the differences between two simulated complex impedance values or between a simulated complex impedance value and a measured value. The FoM will yield a single number describing the discrepancies between two complex impedance values. Lets assume that we have two complex impedance values, Z_{m1} and Z_{m2} and we want to compare the two. First we have to make a choice about the reference. Lets assume that we take Z_{m1} as a reference and compare the Z_{m2} value to Z_{m1}. We now assume that we have a black box with input impedance Z_{m1}^*, see Fig. 1.8.

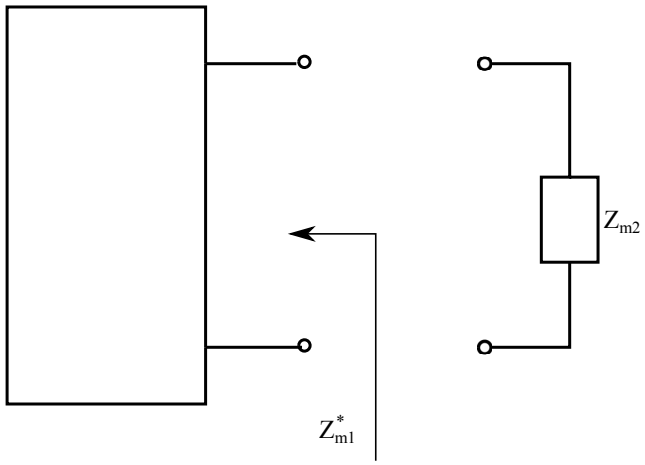


Figure 1.8. 'Connecting' the impedance to be evaluated to a black box having an input impedance that is the complex conjugate value of the reference impedance.

If we 'connect' an impedance Z_{m1} to this black box, we will have a perfect impedance match and the power wave reflection coefficient for this imaginary system, our Figure

of Merit, will be zero. We are not connecting Z_{m1} to the imaginary black box however but Z_{m2}. The amplitude of the power-wave reflection coefficient, our FoM, will give us a single number describing how 'far off' this complex impedance is from the desired value Z_{m1}. Moreover, the FoM relates to real physics, describing how the mismatch will relate to the reflection of power. A reflection of 10 % of the input power (FoM$\leq$ -10 dB) is considered as acceptable and realizable in practice.

Chapter two

Rectifier Analysis for RF Energy Harvesting and Power Transport

In this chapter a complete design method to accurately calculate the impedance and the DC output voltage of the rectifier part in a wireless power transmission system is presented. The presented method uses circuit analysis to calculate the input impedance of a Schottky diode as a function of frequency for different RF input power levels. For the DC output voltage, the Ritz-Galerkin method is used to predict the output DC voltage across different load resistances. Moreover, the influence of the diode parameters on the output voltage is discussed. All the simulated results are validated by measurements and by Harmonic Balance (HB) simulations.
The rectenna design criteria with respect to the operating frequency and the input power level are also motivated in this chapter.

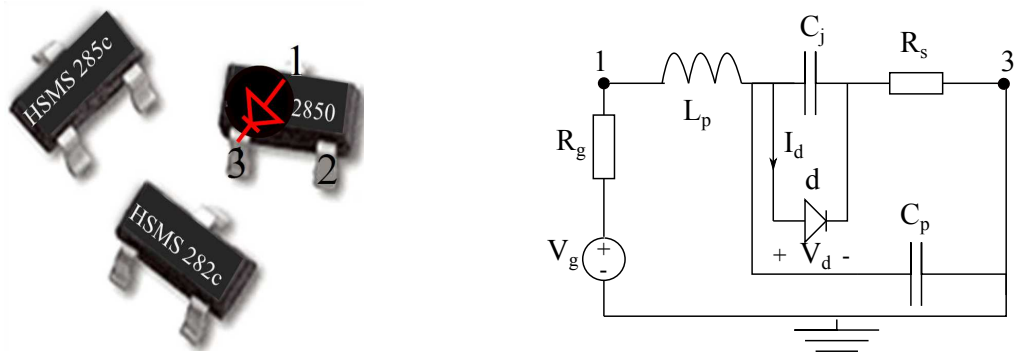
2.1 Introduction

Energy harvesting from renewable sources such as solar light and wind, as well as from other sources like heat and vibration, have been the focus of research recently [7]. However, an increasing number of applications, especially inside buildings where light/solar energy is not always sufficient, cannot profit from these sources. Power transport [35] might be a perfect substitute for batteries to drive wireless sensor nodes.
Recent achievements have shown that far-field energy harvesting [3] and power transport, i.e. means to send power over distances, may overcome the lack of different power sources such as light, heat and vibration.
The basic far-field energy harvester consists of an antenna, which captures part of the transmitted RF signal, connected to a rectifier (AC/DC converter) which converts the AC captured signal into a DC power signal. A power management circuit is needed to guide and store the DC power. For maximum power transfer, a matching network is needed between the rectifier and the antenna [36, 37]. The design, fabrication and measurement of an RF harvester employing a matching network will be presented in Chapter 3.
In order to decrease the size of the system, the matching network is removed in [38] and [39] and the antenna is directly matched to the rectifier ($Z_{antenna} = Z^*_{rectifier}$). The design, fabrication and measurement of a rectenna directly matched to the rectifier circuit will be presented in Chapter 5. In both situations, i.e. with or without a matching network, an accurate analysis of the rectifier's RF input impedance is crucial. This chapter investigates the input impedance of the rectifier as a function of frequency for different power levels. In addition, the influence of the diode's parameters on the output voltage and consequently on the rectifier efficiency are discussed.

2.2 Schottky Diode Equivalent Circuit Model

The diode analysis method to be presented has been introduced in [40] and verified in [33, 41, 42]. The method uses circuit analysis to calculate the input impedance of a packaged Schottky diode as a function of frequency for different power levels.

Figure 2.1(a) shows some commercially available Schottky diodes in SOT-23 package. The equivalent electrical circuit is shown in Fig. 2.1(b). C_j is the junction capacitance, C_p and L_p are the parasitic packaging capacitance and inductance respectively. d is the ideal diode. V_g is an RF voltage generator and R_g is the internal resistance of the generator. Since the Schottky diode is a non-linear device, and its input impedance changes as a function of input power and frequency, the main objective is to present a standard analysis that is fast, accurate and reliable to calculate the impedance of the Schottky diodes.



(a) (b)

Figure 2.1. Commercially available, packaged Schottky diodes (SOT-23) (a) and their equivalent circuit model (b).

To be able to determine the impedance of the rectifier between pins 1 and 3 (see Fig. 2.1), the voltage across and the current through the diode d should be determined. The electrical behavior of the rectifier is investigated assuming a voltage source with a single frequency f_0 as input, $V_g = |V_g| \cos(2\pi f_0 t)$. The electrical behavior of this circuit can be described with the following expressions, obtained by applying Kirchhoff's relations:

$$V_g = I_g R_g + L_p \frac{\partial I_g}{\partial t} + V_{C_P}, \tag{2.1}$$

$$V_{C_P} = V_d + V_{R_s}, \tag{2.2}$$

$$V_{R_S} = R_s(I_{C_j} + I_d), \tag{2.3}$$

$$I_{C_j} = Cj \frac{\partial V_d}{\partial t}, \tag{2.4}$$

$$I_d = I_s(e^{\alpha V_d} - 1), \tag{2.5}$$

so that

$$\frac{\partial V_d}{\partial t} = \frac{1}{R_s C_j} \left\{ \psi \left(\frac{\partial I_g}{\partial t} \right) - R_s I_s \left(e^{\alpha V_d} - 1 \right) \right\}, \tag{2.6}$$

where $\psi \left(\frac{\partial I_g}{\partial t} \right) = V_g - R_g I_g - V_d - L_p \frac{\partial I_g}{\partial t}$. $\alpha = \frac{q}{nKT}$, where $\frac{q}{KT}$ is the thermal voltage and n is the ideality factor. With the aid of the fourth-order Runge-Kutta method (RK4), Eq. (2.6) can been solved as in [33], and the voltage V_d across the diode d is calculated. By using Eq. (2.5), the current I_d flowing through the diode is determined. After the evaluation of V_d and I_d the input impedance of the diode d is found using Ohm's law:

$$Z_d = \frac{V_d}{I_d}. \tag{2.7}$$

After evaluating the input impedance of the diode Z_d, using the (Spice) parameters extracted from the manufacturer data sheet, the input impedance of the Schottky diode Z_{in} including its series resistance and packaging parasitics is calculated. Table 2.1 summarizes the Spice parameters for two commercially available Schottky diodes, Avago HSMS-2820 and HSMS-2850 [43]. C_{23} accounts for a second diode between pins 2 and 3 and is determined empirically from measurements.

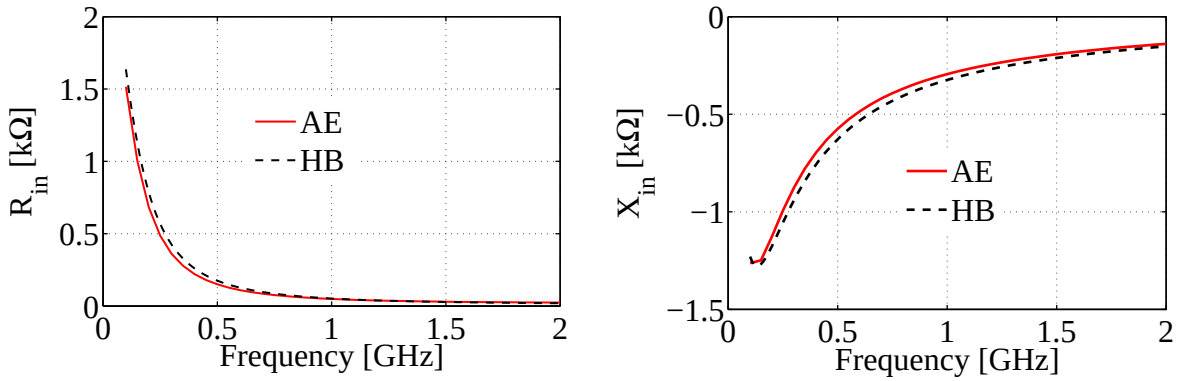
To verify the accuracy of the equivalent circuit model, the two different Schottky diodes HSMS-2820 and HSMS-2852 will be investigated. The input impedance versus frequency for different power levels is calculated with the aid of the equivalent circuit model and is compared with harmonic balance simulation results [44] and with measurement results.

Table 2.1. Spice parameters used to calculate the impedance of the rectifier shown in Fig. 2.1(b) using the analytical procedure.

Parameter	**HSMS-2820**	**HSMS-2850**	Units
I_s	2.2E-8	3.0 E-6	A
n	1.08	1.06	-
R_s	6.0	25	Ω
C_j	0.7	0.18	pF
L_p	2.0	2.0	nH
C_p	0.08	0.08	pF
C_{23}	0.06	0.06	pF

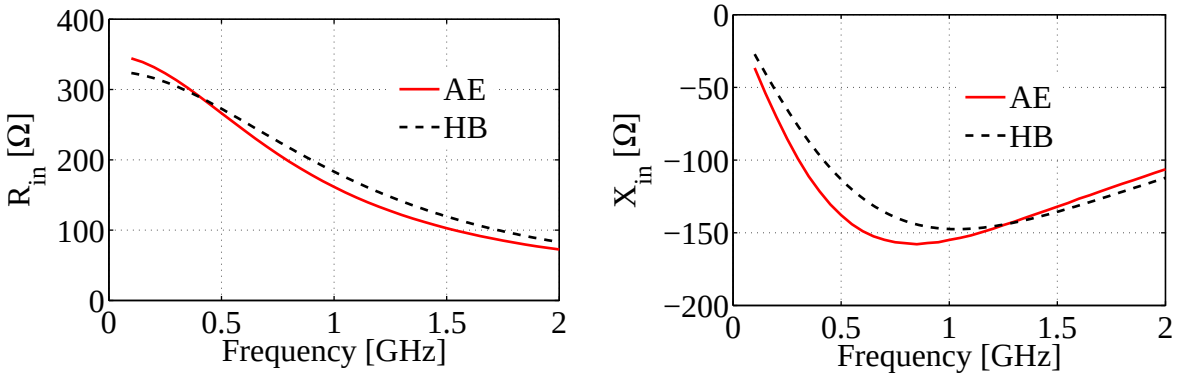
Figures 2.2 and 2.3 show the impedance of the rectifier as calculated by using ADS harmonic balance simulations and the presented analytical expressions. Figures 2.2(a) and 2.2(b) show the real and imaginary parts of the input impedance versus frequency at an input power level of -20 dBm. It is shown in the figures that the equivalent circuit model can predict the impedance of the rectifier with a relative difference of less than 10 % for the real and imaginary parts of the input impedance. In the next paragraph, we will investigate whether this accuracy is sufficient.

In Figs. 2.3(a) and 2.3(b) the input power level is increased to -10 dBm. The input impedance is again calculated using ADS harmonic balance simulations and the equivalent circuit model. It is shown in the figure that the results of the analytical equations are matching the results obtained using harmonic balance simulations. The relative difference for the input impedance calculated by Harmonic Balance (HB) simulation and by the Analytical Expressions (AE) will be presented in detail next.



(a) (b)

Figure 2.2. Real (a) and imaginary (b) parts of the input impedance of the Schottky diode HSMS-2852 versus frequency for a maximum available power level of -20 dBm calculated using the analytical equations (AE) and harmonic balance (HB) simulations. $R_g = 50\Omega$.



(a) (b)

Figure 2.3. Real (a) and imaginary (b) parts of the input impedance of the Schottky diode HSMS-2852 as a function of frequency for a maximum available power level of -10 dBm calculated using the analytical equations (AE) and harmonic balance (HB) simulations. $R_g = 50\Omega$.

To quantify the accuracy of the simulated results shown in Fig. 2.2 and in Fig. 2.3, the Figure of Merit introduced in the previous chapter (see Sec. 1.6), is used. Here, $Z_{m_1} = Z_{AE}$ and $Z_{m_2} = Z_{HB}$, with $Z_{AE} = R_{AE} + jX_{AE}$ and $Z_{HB} = R_{HB} + jX_{HB}$. R_{AE} and X_{AE} are the real and imaginary parts calculated using the presented equivalent circuit model. R_{HB} and X_{HB} are the real and imaginary parts of the input impedance calculated using Harmonic Balance simulations. Figure 2.4 shows that the deviations between the results of the analytical circuit model (AE) and the results of Harmonic Balance (HB) simulations do not exceed -10 dB, which satisfies the FoM criterion, which leads to the conclusion that AE results are a good alternalive for HB simulations.

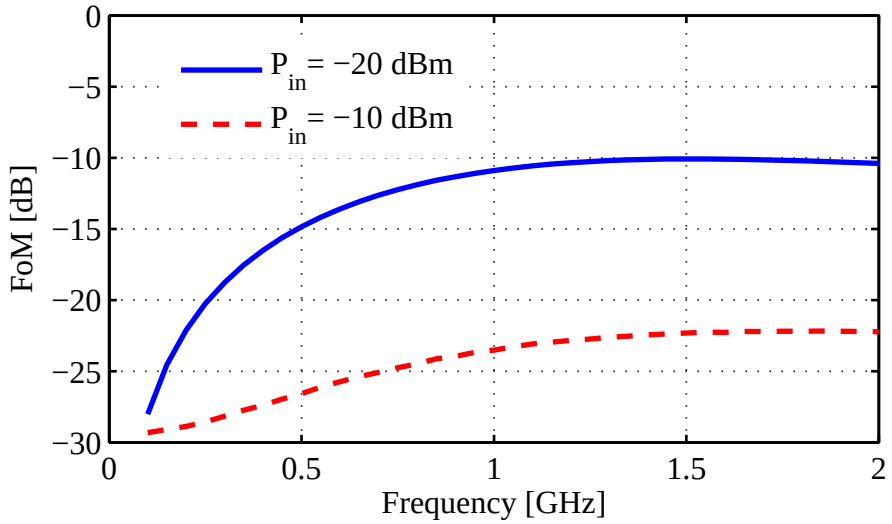
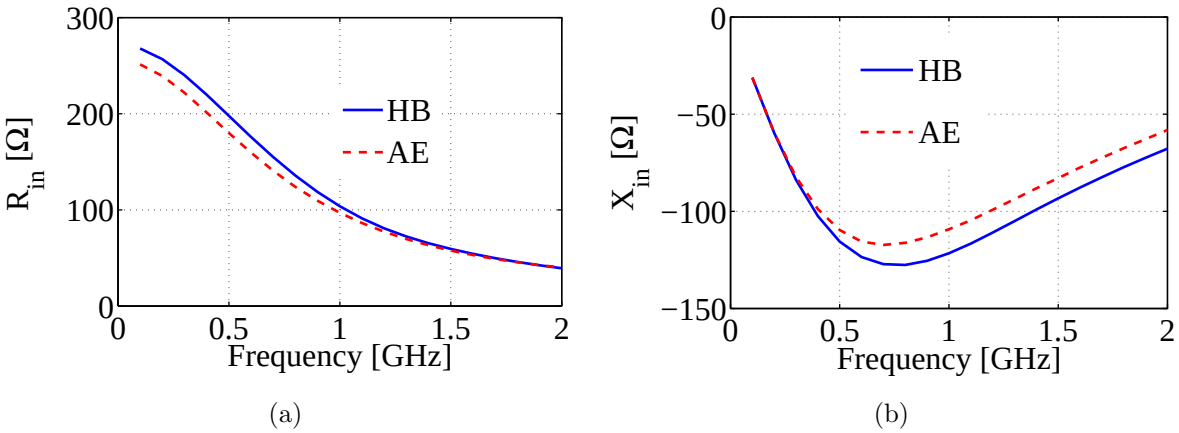


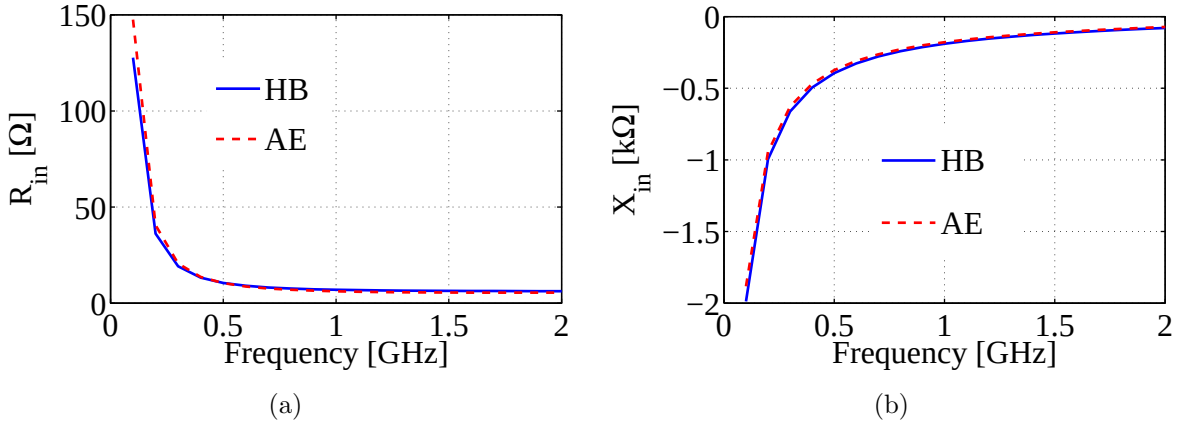
Figure 2.4. Figure of Merit to calculate the deviation between the results of the presented equivalent circuit model (AE) and harmonic balance simulation results (HB) as presented in Figs. 2.2 and 2.3. Note the different vertical axis scales in Figs. 2.2 and 2.3.

The real and imaginary parts of the input impedance of the commercially available Schottky diode Avago HSMS-2820 are also calculated from the analytical expressions presented in this section and are compared to the ones obtained by Harmonic Balance simulations [44]. The results at an input power level of 0 dBm and -10 dBm are plotted in Fig. 2.5 and Fig. 2.6, respectively.



(a) (b)

Figure 2.5. Real (a) and imaginary (b) parts of the input impedance of the commercially available Schottky diode Avago HSMS-2820 versus frequency for a maximum available power level of 0 dBm calculated using the analytical expressions and harmonic balance simulations. The source has a 50Ω internal resistance.



(a) (b)

Figure 2.6. Real (a) and imaginary (b) parts of the input impedance of the commercially available Schottky diode Avago HSMS-2820 versus frequency for a maximum available power level of -10 dBm calculated using the analytical expressions and harmonic balance simulations. The source has a 50Ω internal resistance.

The deviations between AE and HB have been calculated as a function of frequency using the Figure of Merit introduced in Sec. 1.6. The deviations do not exceed -10 dB, which validates the accuracy of the presented equivalent circuit model, taking the Harmonic Balance results as a reference. What remains is to confirm the correctness of taking the Harmonic Balance simulation results as a reference. Therefore, we will compare both simulations (AE and HB) with measurements.

2.3 Measurement Setup and Measured Results

In this section, the equivalent circuit model will be verified by measurements. In order to fully verify the accuracy of the presented analytical equations, the commercial Schottky diode HSMS-2822 is used for measurements. To measure the impedance of the rectifier the setup shown in Fig. 2.7 is used. The Device Under Test (DUT) is mounted directly on a SMA connector and then connected to the Vector Network Analyzer (VNA) using a phase-stable coaxial cable. A DC blocker is used to prevent DC currents flowing through the VNA. After calibrating, the phase reference point is at the SMA connector (point 1 in Fig. 2.7), so that the phase-stable coaxial cable of length l has no influence on the measurements. The PNA-X Network Analyzer [45] is used to measure the reflection coefficient S_{11} at the DUT (Voltage reflected due to the mismatch between Z_0 and Z_{DUT}).

$$S_{11} = \frac{Z_{DUT} - Z_0}{Z_{DUT} + Z_0}. \tag{2.8}$$

The input impedance of the rectifier $Z_{rectifier}$ is calculated from the reflection coefficient

$$Z_{rectifier} = Z_0 \frac{1 + S_{11}}{1 - S_{11}}, \tag{2.9}$$

where $Z_0 = 50\Omega$. It should be mentioned that this measurement setup can only be used to measure the impedance of two or more diodes in parallel. If the impedance of a single diode has to be measured, the measurement setup in Fig. 2.7 should be modified and an inductor should be connected in parallel to the rectifier so that the diode can self-bias itself allowing the diode's DC current to return to ground. Pins 1,2 and 3 in Fig. 2.7 correspond to the ones indicated in Fig. 2.1(a).

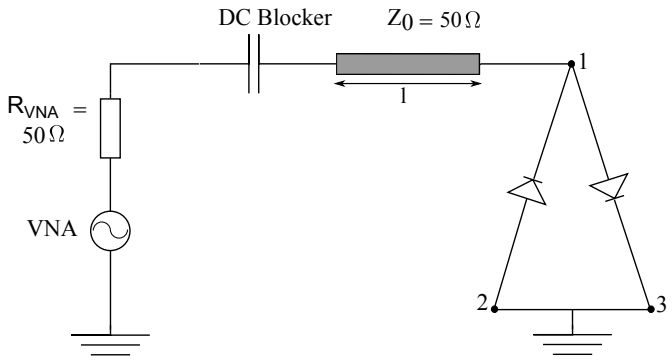
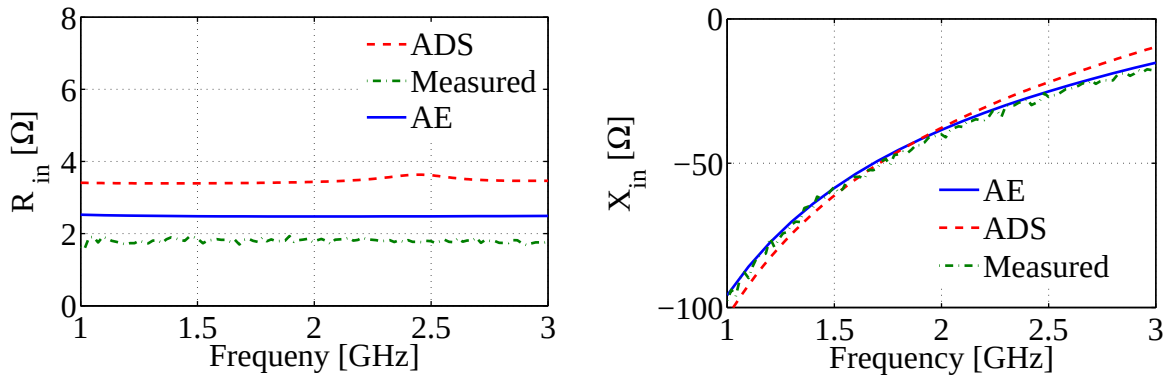


Figure 2.7. Setup to measure the impedance of the diode using a Vector Network Analyzer (VNA).

Figures 2.8(a) and 2.8(b) show the real and imaginary parts of the input impedance as a function of frequency at an available power level of −15 dBm.



(a) (b)

Figure 2.8. Real (a) and imaginary (b) parts of the input impedance versus frequency for the diode HSMS-2822 at an available power level of −15 dBm.

Figures 2.9(a) and 2.9(b) show the input impedance as a function of available power level P_{in} at an operating frequency of 2.45 GHz.

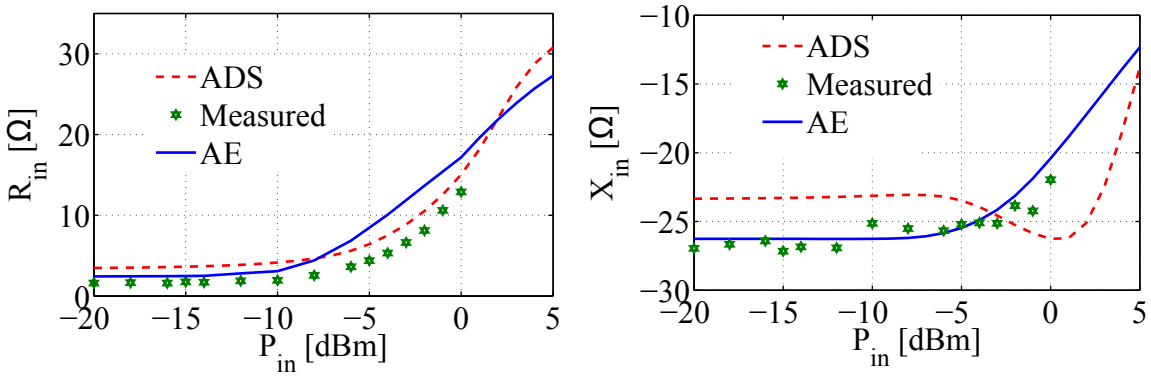


Figure 2.9. Real (a) and imaginary (b) parts of the input impedance versus P_{in} for diode HSMS-2822 at an operating frequency of 2.45 GHz.

It is demonstrated in Figs. 2.8 and 2.9 that the presented analytical expressions as well as Harmonic Balance simulations can predict the impedance behaviour of the diode as a function of frequency (see Fig. 2.8) and as a function of the available power level (see Fig. 2.9). To quantify the agreement between the results of the AE and measurements, the Figure of Merit as a function of input power level is plotted in Fig. 2.10. It is depicted in the figure that the deviations do not exceed the -10 dB threshold, thus validating the accuracy of the analytical expressions with respect to measurements.

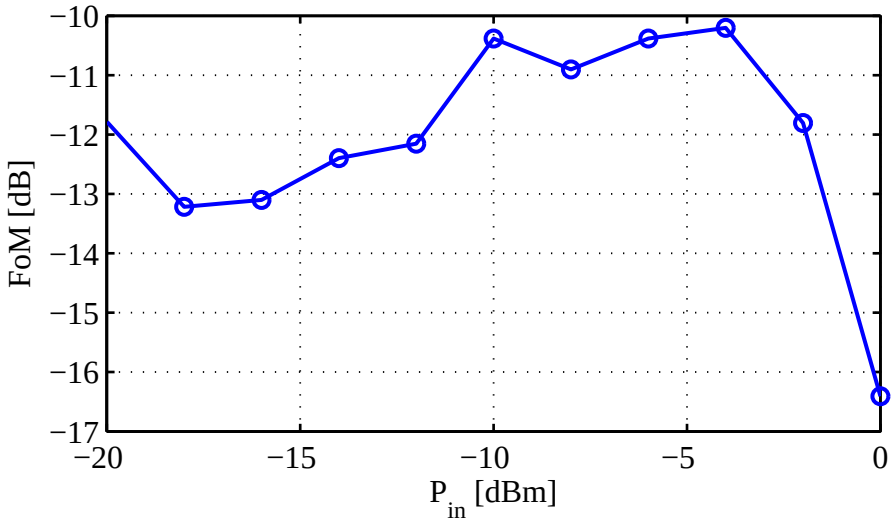


Figure 2.10. Figure of Merit versus maximum available power level for the equivalent circuit model and the measurements shown in Fig. 2.9.

2.4 Output Voltage of the Rectifier Circuit

In this section we will investigate the DC output voltage as a function of maximum available input power for a single diode. The Ritz-Galerkin averaging method [46] is used to predict the DC output voltage across different resistance values. An implicit equation is obtained for the output voltage as a function of the maximum available input power P_{in}, the generator resistance R_g, the saturation current I_s, the contact resistance R_s and the load resistance R_L,

$$\mathrm{I}_0\left(\alpha\sqrt{8R_gP_{in}}\right) = \left(1+\frac{V_{out}}{R_LI_s}\right)\exp\left(1+\frac{R_g+R_s}{R_L}\right)\alpha V_{out} \tag{2.10}$$

where $\mathrm{I}_0\left(\alpha\sqrt{8R_gP_{in}}\right)$ is the zero-order modified Bessel function of the first kind. $\alpha = \frac{q}{nkT}$, where q is the electron charge, k is Boltzmann's constant, T is the temperature in Kelvins, and n is the diode ideality factor. By solving Eq. (2.10) the output voltage for different load resistances is calculated. Note that the generator is not matched to the rectifier and that P_{in} is the power that would be available if the diode would have the same impedance as that of the generator. Therefore, P_{in} is the maximum available power. The output voltage of the commercially available diode HSMS-2850 ($I_s = 3\,\mu$A and $R_s = 25\Omega$, $n = 1.03$) is calculated and compared to harmonic balance simulation results. Figure 2.11 shows the DC output voltage calculated by the Ritz-Galerkin (RG) method and by Harmonic Balance (HB) simulations for different load resistances.

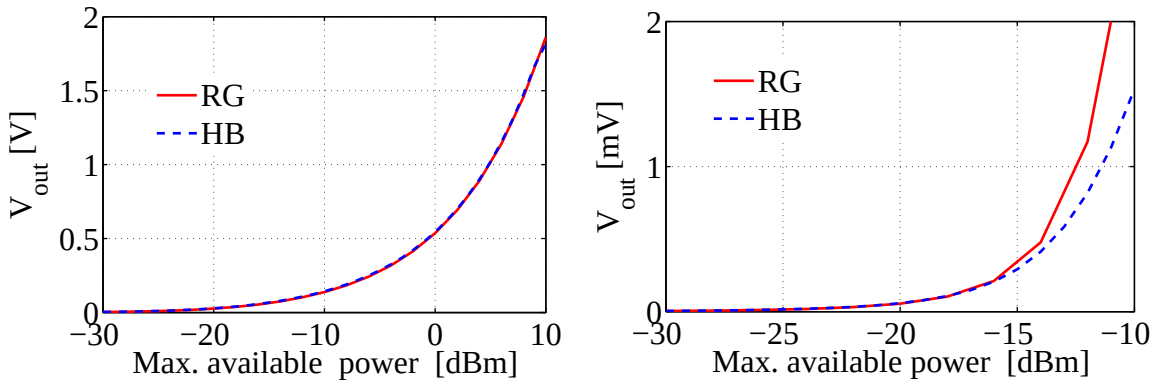


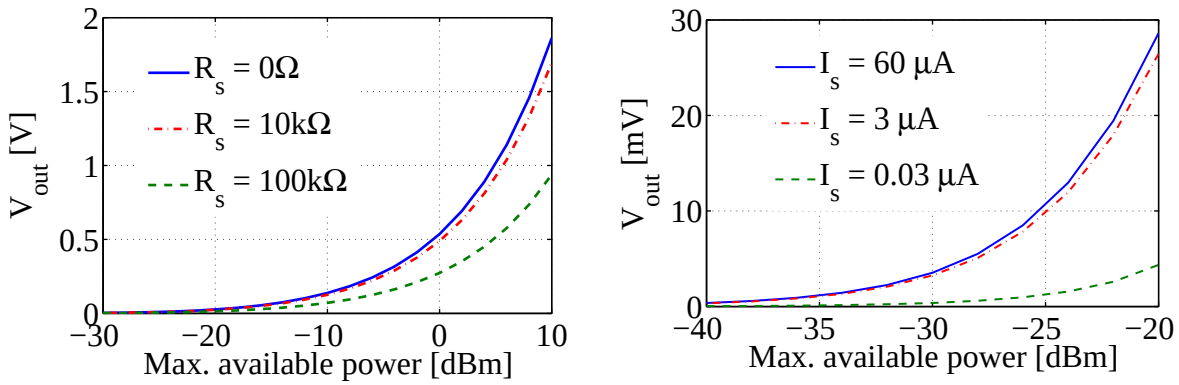
Figure 2.11. Output voltage as a function of maximum available input power calculated by Ritz-Galerkin averaging (RG) method and by Harmonic Balance (HB) for a load resistance of (a) $R_L = 100\mathrm{k}\Omega$, and (b) $R_L = 10\Omega$. The generator impedance $R_g = 50\Omega$.

It is shown in the figure that the simulation results calculated by Ritz-Galerkin aver-

aging method can accurately predict the output voltage of the rectifier with a relative difference of less then 5 %, when the load resistance is high (100kΩ). On the other hand when the load resistance decreases the relative difference increases to 43 % for $P_{in} = -12$ dBm. For a 10 Ω load resistance, the Ritz-Galerkin averaging method is accurate up to Pin = -15 dBm (relative difference $\leq 5\%$) as depicted in Fig. 2.11(b).

2.5 Investigation of the Diode Parameters on the Output Voltage

The influence of the diode parameters (the junction capacitance C_j and the contact resistance R_s) on the conversion efficiency was first investigated in [47]. In this section we investigate the effect of the contact resistance R_s and the saturation current I_s on the output voltage. The Ritz-Galerkin averaging method [46] presented in the previous section is used to calculate the DC output voltage as a function of maximum available input power level for different contact resistance values R_s and for different saturation current values I_s. Figure 2.12(a) shows the output voltage of the rectifier as a function of maximum available power for different values of the contact resistance R_s. Figure 2.12(b) shows the output voltage versus maximum available power for different saturation current values I_s. The load resistance R_L is set to 100 kΩ.
The figure shows that the DC output voltage increases with a decreasing value of R_s and increases with an increasing value of I_s. These results provide additional selection criteria for choosing commercially available Schottky diodes.



(a) (b)

Figure 2.12. Influence of the contact resistance (a) and the saturation current (b) on the output voltage as a function of maximum available power. $R_L = 100$ kΩ.

Table 2.2 lists the I_s and the R_s values of different commercially available Schottky

diodes.

Table 2.2. I_s and R_s values for commercially available Schottky diodes.

Schottky diode	I_s (A)	R_s (Ω)
HSCH-5310	3.00E-10	13
HSCH-5314	3.00E-10	9
HSCH-5340	4.00E-8	13
HBAT-540X	1.00E-7	2.4
HSMS-270X	1.40E-7	0.65
HSMS-280X	3.00E-8	30
HSMS-281X	4.80E-9	10
HSMS-282X	2.20E-8	6
HSMS-285X	3.00E-6	25
HSMS-286X	5.00E-8	6

Figure 2.13 shows the calculated DC output voltage as a function of the load resistance (R_L) for two different maximum available power levels, for three chosen Schottky diodes. The figure shows that the Schottky diode HSMS-285X outperforms the two other diodes HSCH-5310 and HSMS-280X with respect to the DC output voltage for the two simulated maximum available power levels. This is due to the higher saturation current.

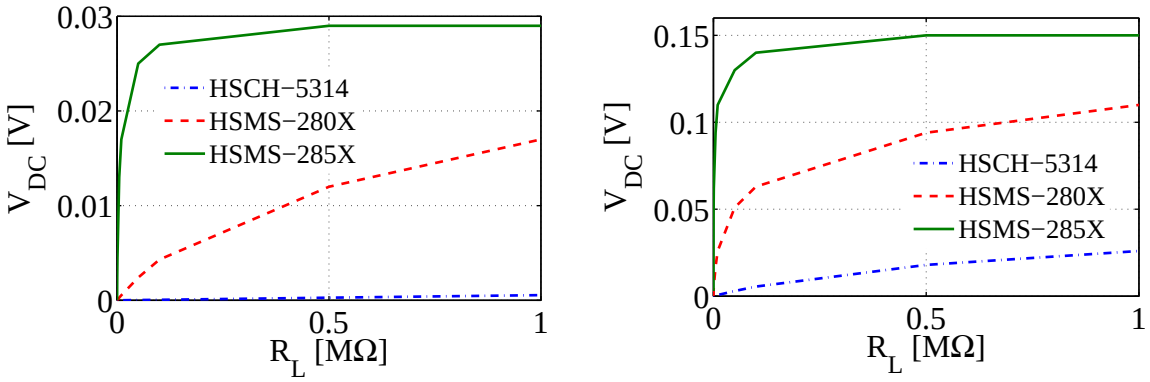


Figure 2.13. DC output voltage as a function of load resistance (R_L) for different maximum available power levels for three Schottky diodes. (a) $P_{in} = -20$ dBm, (b) $P_{in} = -10$ dBm.

The Schottky diode HSMS-285X is characterized by a saturation current that is higher

than that of the other diodes. Since this leads to a higher DC output voltage, this diode is selected for rectenna designs to be discussed in subsequent Chapters.

2.6 Voltage Multiplier

At low input power levels, the DC output voltage of the rectifier is well below 1 V. Modern low-power electronics needs at least a voltage in the order of 1 - 1.2 V to operate [48]. To increase the DC output voltage, a voltage multiplier may be employed. A much used configuration is based on a Dickson charge pump circuit [49] and is shown in Fig. 2.14.

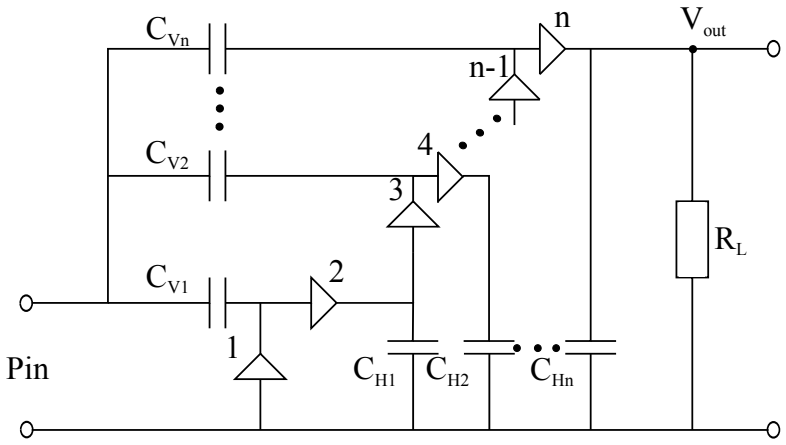


Figure 2.14. Voltage multiplier configuration based on a Dickson charge pump circuit.

As an illustration, ADS harmonic balance has been used to simulate the DC output voltage. For an operating frequency of 1 GHz and based on the Schottky diode HSMS 2822, the DC output voltage across R_L has been simulated as a function of the number of rectifier stages. The results are shown in Fig. 2.15 for two maximum available power levels.

At an input power level of +10 dBm (solid curve), the maximum output voltage is reached after cascading 8 Schottky diodes, while at a power level of −6dBm (dashed curve), the output voltage is saturated after cascading only 4 Schottky diodes. It is clearly shown in Fig. 2.15 that, at a low input power level (e.g. -6 dBm), increasing the number of the rectification stages will eventually result in a lower DC output voltage, and consequently the RF-to-DC power conversion efficiency will decrease.

The saturation point depends on the operating frequency, the input power level, and the load resistance value. In view of this saturation effect, and since this thesis deals with harvesting power at very low input power levels ($P_{in} = -10$ dBm) a voltage doubler configuration (n=2, see Fig. 2.14) will be adapted for the designed rectennas in Chapter

3 and in Chapter 5.

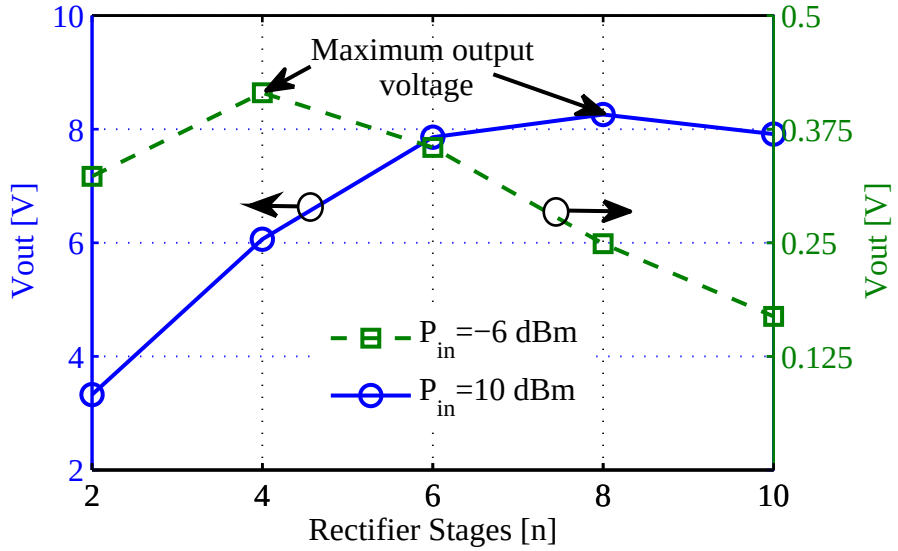


Figure 2.15. DC output voltage as a function of the number of multiplier stages (diodes) for two maximum available power levels simulated by ADS. Frequency = 1 GHz, R_L =1MΩ, $C_{Vi} = C_{Hj}$ = 100 pF. R_g = 50Ω. P_{in} is the maximum available input power.

After choosing the commercially available Schottky diode HSMS-285X for the design of the rectenna and limiting the multiplier to a voltage doubler, the following section will present the selection of the design frequency and the input power level.

2.7 Rectenna Design

Since the rectifier's impedance changes as a function of frequency and as a function of input power level, an operating frequency f_0 and an input power level (P_r) should be selected first, before proceeding to the investigation of different rectenna topologies. Based on the two design criteria, operating frequency and assessed input power level (f_0, P_r), different rectenna topologies will be presented. Figure 2.16 shows a general far-field wireless power transmission system. The rectenna in this system consists of an antenna that is directly complex-conjugate matched to the rectifier.

The power density (S) at the receiving side is dictated by the effective isotropic radiated power, $EIRP = P_t G_t$, where P_t is the transmitted power and G_t is the transmit antenna gain, and by the distance R between the transmitting and the receiving antennas as shown in Fig. 2.16.

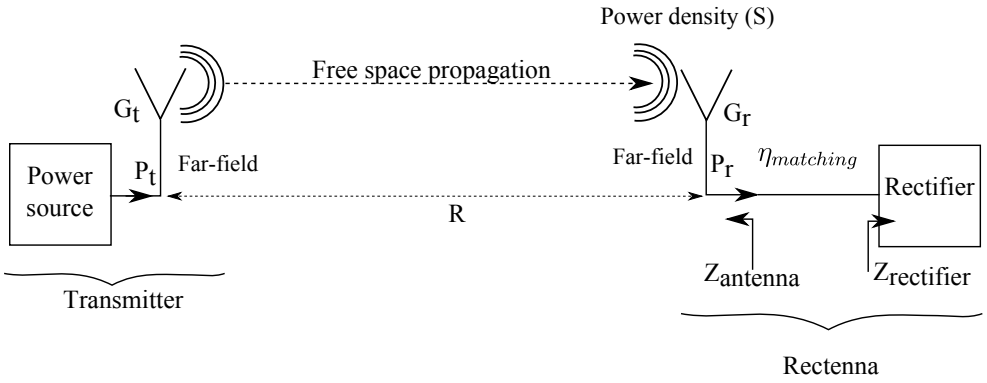


Figure 2.16. Far-field wireless power transmission system, where no matching circuit is used at the receiving side.

The power density is obtained by distributing the total power over a spherical surface with area $4\pi R^2$:

$$S = \frac{EIRP}{4\pi R^2} = \frac{P_t G_t}{4\pi R^2}. \tag{2.11}$$

The antenna at the receiving side captures part of this power density with an effective aperture A_r. The received power P_r (see Fig. 2.16) is given by:

$$P_r = A_r S, \tag{2.12}$$

where A_r is the effective aperture of the receiving antenna in the direction of the transmitting antenna. According to [50], the effective aperture is related to the antenna gain by

$$A_r = G_r \frac{\lambda_0^2}{4\pi}, \tag{2.13}$$

where λ_0 is the wavelength of transmitted signal. Substituting Eq. (2.13) in Eq. (2.12) leads to the Friis transmission equation where the received power at the terminals of the rectifier is expressed as

$$P_r = P_t G_t G_r \left(\frac{\lambda_0}{4\pi R}\right)^2 = P_t G_t G_r \left(\frac{c}{4\pi f_o R}\right)^2, \tag{2.14}$$

where c is the free-space speed of light and f_0 is the operating frequency of the transmitted signal. The term $\left(\frac{c}{4\pi f_o R}\right)^2$ is called the free-space path loss. The term path loss may lead to confusion. It gives the impression that, apart from a spherical spreading, a frequency dependent path loss exists. This is not true, since the gain changes with frequency.

2.7.1 Operating Frequency

For practical reasons we will aim for a rectenna that is ideally credit-card sized, so that it can be applied in wireless sensors for smart building applications.

Since there are no special regulations for rectenna systems yet, it makes sense to operate these systems in the licence-free Industry-Science-Medical (ISM) frequency bands, see Table 2.3. The frequency bands around 0.9 GHz and 2.4 GHz are of special interest due to our credit card size constraint and due to the relative high EIRP allowed. Adapting a frequency band from 865 to 928 MHz will allow RF power transfer operation both in Europe and North America, and RF harvesting from GSM transmissions on both continents.
After having thus selected the design frequency, the expected input power level will be presented next.

Table 2.3. Frequency band allocations and transmit power restrictions for RFID applications as reported in [12].

Frequency Band	Power	Duty Cycle/ Tx type	Channel Spacing / BW	Region
2446 - 2454 MHz	500 mW EIRP 4 W EIRP	Up to 100 $\leq$ 15%	No spacing	Europe*
865.0 - 865.6 MHz	10 mW ERP		200 kHz	Europe
865.6 - 867.6 MHz	2 W ERP		200 kHz	Europe
867.6 - 868.0 MHz	500 mW ERP		200 kHz	Europe
902 - 928 MHz	4 W EIRP	FH ($\geq$ 50 channels) or DSSS		USA Canada
2400-2483.5 MHz	4 W EIRP	FH ($\geq$ 75 channels) or DSSS		USA Canada
2400-2483.5 MHz	10 mW EIRP		1MHz BW	Japan Korea

In Table 2.3, ERP is the Efective Radiated Power, the EIRP is 1.64 times ERP. FH stands for Frequency Hopping. DSSS stand for Direct Sequence Spread Spectrum and BW indicates Band Width. *Power levels above 500 mW are restricted to use inside the boundaries of a building and the duty cycle of all transmissions shall in this case be $\leq$ 15% in any 200 ms period (30 ms on / 170 ms off).

2.7.2 Expected Input Power Level

Possible design operating frequencies have been discussed in the previous section. We will settle for a design frequency of 868 MHz. It is indicated in Table 2.3 that at this design frequency, the maximum allowed Effective Radiated Power (ERP) is 500 mW in Europe.
Based on the applications that have been indicated in the previous chapter, including the Smart Building Initiative (SBI), a direct line-of-sight propagation distance (R) between 2 meters and 10 meters is expected, being related to the size of a general office, hallway or meeting room. Using the maximum allowed ERP at a frequency of 868 MHz, see Table 2.3, the power density as a function of propagation distance is calculated using Eq. (2.11) and is plotted in Fig. 2.17 (solid curve). It is indicated in the figure that the power density does not exceed 12 dBm/m^2. At a distance of 5 meters away from the transmitting source, the power density drops to 4 dBm/m^2.

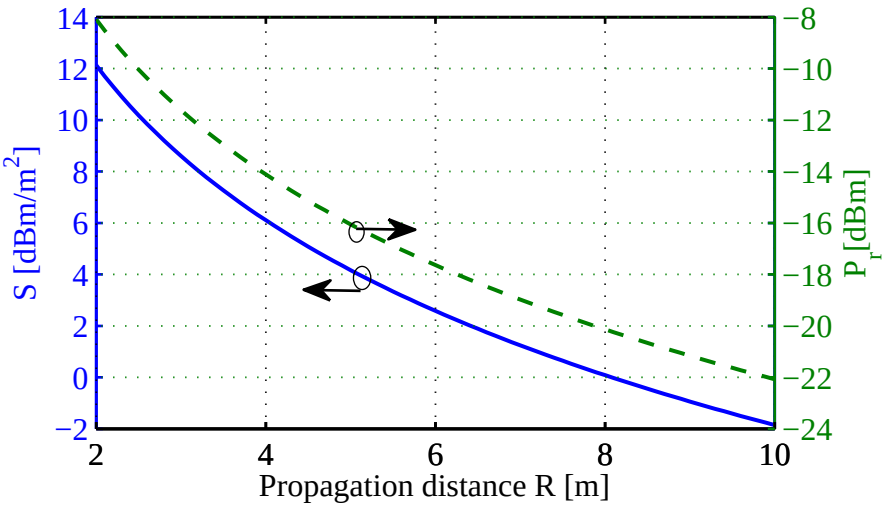


Figure 2.17. Power density and received power calculated using Eq. (2.11) and Eq. (2.14) respectively as a function of propagation distance.

To translate the power density into input power levels P_r (see Fig. 2.16), Eq. (2.14) is used to calculate the input power level, where a receiving antenna gain of 0 dBi[1] is used. The received power within the practical propagation distance ($2 \leq R \leq 10$ m) is plotted in Fig. 2.17 (dashed curve). It is clearly indicated in the figure that the received power is within -8 dBm $\leq P_r \leq -22$ dBm for the practical propagation distances. Based on this argument it makes sense then to choose an assessed input power level of -10 dBm.

[1]The reason behind choosing 0 dBi for the receive antenna gain is the size constraint related to a credit-card size antenna at 868 MHz.

A second reason to choose an input power level around -10 dBm is derived from the results of the input impedance of the Schottky diode as a function of maximum available input power level shown in Fig. 2.9. Since the input impedance of the Schottky diode (both the real and the imaginary part) is almost constant at low input power levels ≤ -5 dBm, choosing a low input power level for the design of the rectenna will ensure that the rectenna will be matched for a broad input power-level band.

To further strengthen this power level choice we will look at the antenna matching. To this end, the term matching efficiency denoted by $\eta_{matching}$ is introduced. The matching efficiency $\eta_{matching}$ describes how well the antenna is matched to the rectifier. Since a matching network is not present (see Fig. 2.16), the antenna is conjugately matched to the rectifier, with an impedance $Z_{antenna} = Z^*_{rectifier}$, where * denotes the complex conjugate. The matching efficiency is calculated as a function of received power level P_r, for an antenna designed to be matched at $P_r = -10$ dBm and for and antenna designed to be matched at $P_r = 0$ dBm.

The matching efficiency is expressed as

$$\eta_{matching} = 1 - S_{11}^2, \tag{2.15}$$

where the power wave reflection coefficient concept [34] [51] is used to calculate the reflection coefficient:

$$S_{11} = \frac{Z_{rectifier} - Z^*_{antenna}}{Z_{rectifier} + Z_{antenna}}. \tag{2.16}$$

Figure 2.18 shows the matching efficiency as a function of input power level for both antennas.

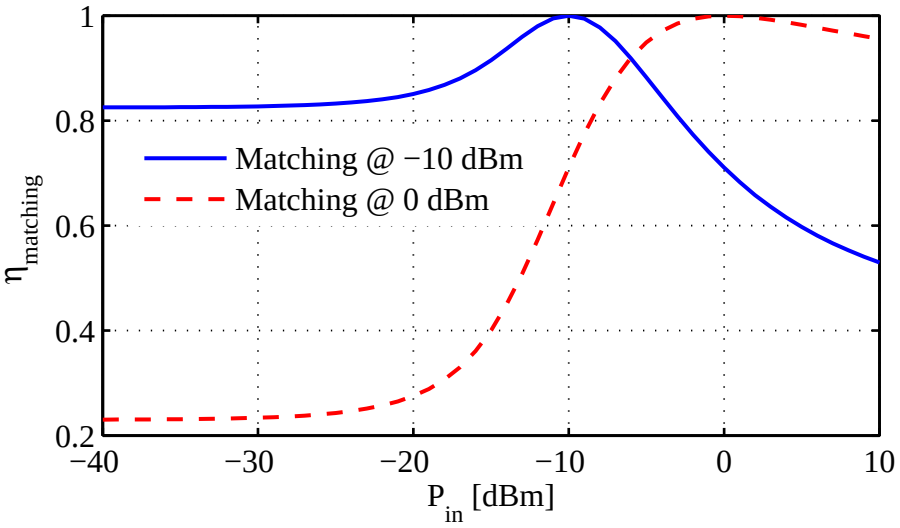


Figure 2.18. Matching efficiency as a function of input power level for an antenna designed to be matched at $P_r = -10$ dBm (solid curve) and for and antenna designed to be matched at $P_r = 0$ dBm (dashed curve).

The figure shows that, when the rectenna is designed for -10 dBm, a matching efficiency higher than 80 % is calculated for a broad input-power-band, $-40\,\text{dBm} \leq P_r \leq -3\,\text{dBm}$. On the other hand, if the rectenna is designed at an input power level of 0 dBm, a matching efficiency higher than 80 % is calculated for an input power band, $-8\,\text{dBm} \leq P_r \leq 10\,\text{dBm}$.
It is thus demonstrated that designing a rectenna for an input power level around -10 dBm will result in a rectenna that is matched to broad input power band from -40 dBm up to -3 dBm, which corresponds to the expected input power levels for our applications (see Fig. 2.17 dashed curve).

2.8 Conclusion

For a maximum power transfer between the receiving antenna and the rectifier in a rectenna system, the antenna's input impedance should be matched to that of the rectifier circuit. A detailed design method to accurately predict the input impedance and the DC output voltage of a Schottky diode based rectifier is presented in this chapter. The presented method uses circuit analysis to calculate the impedance of a Schottky diode as a function of frequency for different maximum available power levels. The simulation results are validated by measurements. In addition, the influence of the diode's parameters on the output voltage are investigated. It is shown that the highest output DC voltage is achieved for the lowest contact resistance and for the highest saturation current. Thus the commercially available Schottky diode HSMS-285X is selected for the design of a voltage doubler.
In addition, the practical rectenna design criteria with respect to the operating frequency and the input power levels are presented. Based on the possible applications, license free frequency bands and allowed transmit powers, chosen is to design rectennas having small or broad frequency bands around 868 MHz for an input power level of -10 dBm ($100\mu\,\text{W}$).

Chapter three

RF Power Transport Employing Rectennas Using Matching Networks

In this chapter a modified Yagi-Uda antenna is introduced. The design is based on a wide-band strip dipole around which a Yagi-Uda structure is formed. The antenna consists of a reflector and a single director placed nearby the driven dipole. The main advantage of the suggested antenna is that, by tuning its geometrical parameters, the antenna can perform as a dual-band, a triple-band or a broad-band antenna. Three prototypes are designed, simulated, fabricated and measured. The simulated results are validated by measurements. The modified Yagi-Uda antenna is then used for RF power transport. A matching network is desigend to match the 50Ω antenna to a voltage-doubler rectifier. The matching circuit and rectifier are characterized as a single unit. The measured Power Conversion Efficiency (PCE) reaches 31.6% and 49.7 % over 10 kΩ load resistance at an input power level of -20 dBm and -10 dBm, respectively. Compared to the state-of-the-art published results, the PCE increases by 5 % at -10 dBm and by 10 % at -20 dBm.
The advantage of this rectenna design is that the antenna may be replaced by any 50Ω antenna having the correct frequency response. The disadvantage is that the matching network introduces additional losses. This disadvantage will be dealt with in a subsequent chapter.

3.1 Introduction

The main challenge in far-field energy harvesting is to overcome the power-density decay due to the long distance between the transmitter and the harvesting location. To partly compensate for this so-called path loss, antenna arrays or high gain antennas can be used at the receiving part. Figure 3.1 shows the 50 Ω rectenna system investigated in this chapter. The arrows indicate the power flow from the transmitting side to the receiving side to reach the load R_L.
First, a novel 50 Ω modified Yagi-Uda antenna for the transmission and reception of RF power is introduced. Three different prototypes are designed, simulated, manufactured and measured. The first antenna is designed to harvest ambient RF power from Digital TV (DTV) broadcasting stations. The second and the third prototype are a dual-band and a triple-band antenna respectively. The dual band prototype is developed for RF power harvesting at 900 MHz (GSM 900) and 2.45 GHz (ISM). The triple-band prototype adds (GSM 1800) to these two frequency bands.

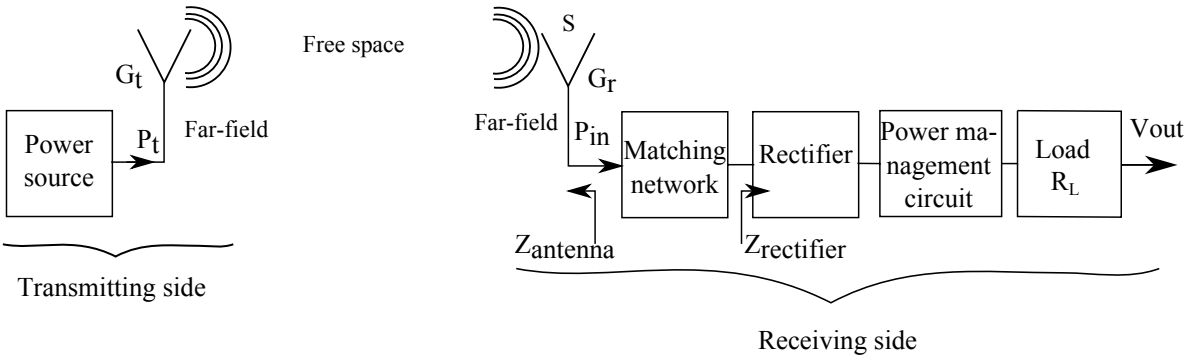


Figure 3.1. RF power transport system employing a (50 Ω) antenna matched to the rectifier using a matching network on the receiving side.

A Schottky diode voltage doubler configuration is used on the receiving side to transform the RF power into DC power. For an improved power transfer, a lumped-element matching network between the 50Ω receiving antenna and the voltage doubler (rectifier) is designed. The main target of the matching network is to transform the impedance of the rectifier ($Z_{rectifier}$) to that of the antenna ($Z_{antenna}$), see Fig. 3.1.

This chapter consists of two parts. In the first part, the design of a novel, modified Yagi-Uda antenna to harvest power from DTV stations is presented. The developed procedure is then employed also to design a dual-band and a triple-band antenna. In the second part, the matching network will be introduced. Using the newly developed antenna, we design one RF power transport prototype system at an operating frequency

of 868 MHz for an input power level of -10 dBm.

For the development of the modified Yagi-Uda antenna, we first need to introduce a wide-band strip dipole antenna that will replace the driven element in a normal Yagi-Uda antenna.

3.2 Center-Fed and Step-Shaped-Fed Strip Dipoles

Various types of printed dipoles have been studied in the literature [52,53]. The printed strip dipole antennas are desirable due to their low profile, low production cost and their ease of fabrication. Many attempts have been performed to enlarge the bandwidth of the strip dipole antenna [54]. A printed wide-band dipole antenna with a step-shaped feed gap for DTV signal reception was suggested in [55] and is shown in Fig. 3.2(b). The wide-band characteristics are obtained by two closely-spaced resonances.

Figure 3.2 shows a conventional centre-fed strip dipole (a) and the step-shaped-fed (the feed of the antenna has a shape of a step strip dipole antenna (b) as introduced in [55] and [56].

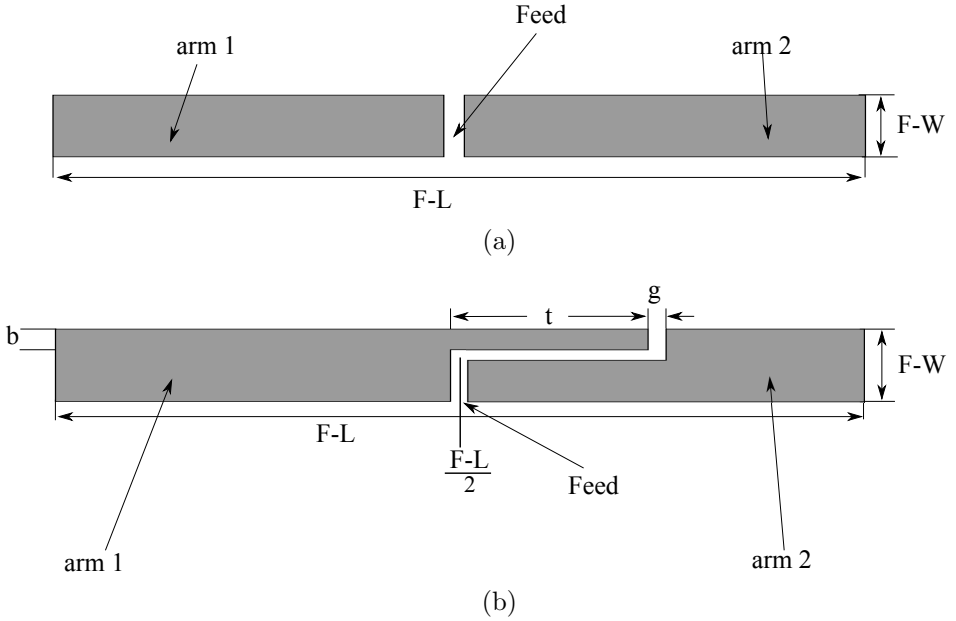


Figure 3.2. Centre-fed strip dipole configuration (a) and the step-shaped-fed strip dipole configuration (b) as introduced in [55].

To verify that the suggested antenna in [55] has two resonance modes, Table 3.1 shows the optimized parameters for having the step-shaped-fed strip dipole resonate at 550

MHz. An ordinary, centre-fed strip, dipole is also designed to resonate at the same frequency. The center-fed and the step-shaped-fed dipole antennas are printed on a 1.6 mm thick FR4 substrate with permittivity $\epsilon_r = 4.4$ and loss tangent $\tan\delta = 0.025$. CST Microwave Studio [57] is used to simulate the reflection coefficients as a function of frequency for both the centre-fed and the step-shaped-fed strip dipole.

Table 3.1. Optimized parameters for the step-shaped-fed dipole to resonate at 550 MHz. The dimensions are explained in Fig. 3.2.

Parameter	mm
F-L	226
F-W	20
b	2.5
g	1
t	42.5

Figure 3.3 shows the simulated reflection coefficient versus frequency for a centre-fed strip dipole antenna having a length F-L and width F-W (see Table 3.1) and for a step-shaped-fed strip dipole antenna of the same length and width. The dimensions of the feed structure of the latter antenna, i.e., b, g and t, are listed in Table 3.1.

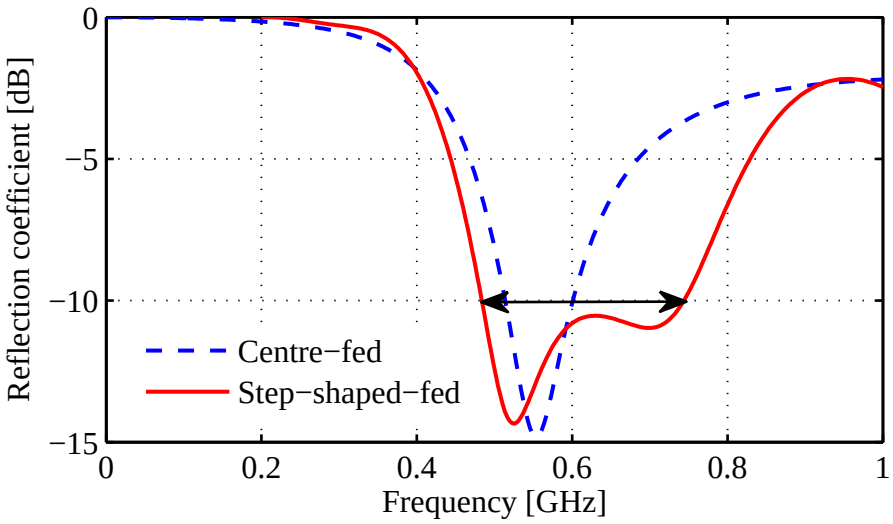


Figure 3.3. Reflection coefficient as a function of frequency for the centre-fed strip dipole antenna and the step-shaped-fed strip dipole antenna, having identical total length and identical width and both placed on the same substrate carrier.

It is indicated in the figure that the bandwidth of the step-shaped-fed strip dipole

antenna is increased by more than 120% compared to the standard centre-fed strip dipole antenna. This increase in bandwidth is realized by the two excited resonant modes, one at 0.5λ controlled by the antenna length F-L, and a second one controlled by choosing a proper value of t (see Fig. 3.2(b)). This additional resonant mode of the antenna is controlled by the total effective length of the antenna comprising its two asymmetric radiators arm 1 and arm 2 [55] (see Fig. 3.2(b)).
The main limitation of this antenna is its restricted gain. Since it is a $\lambda/2$ dipole antenna, its gain is limited to 2.15 dBi. For a higher antenna gain, other types of antennas are needed.

3.3 Printed Yagi-Uda Antenna and Parameter Optimization

The printed Yagi-Uda antenna shown in Fig. 3.4 is a highly directive but narrow-band antenna [58]. The parameters of a quasi Yagi-Uda antenna were optimized for a broad-band operation in [59]. The resulting antenna achieves 48% bandwidth for a Voltage Standing Wave Ratio (VSWR) < 2, but uses a truncated ground plane and a balun structure.

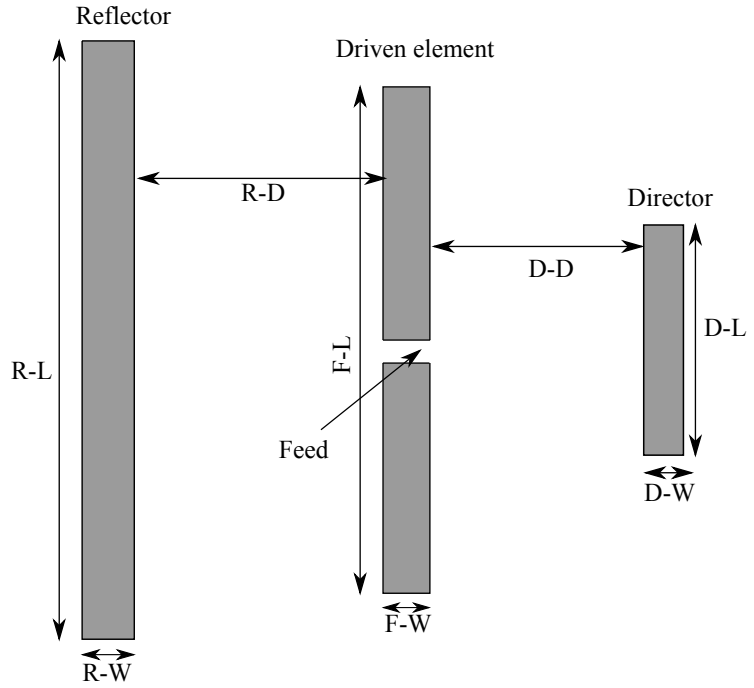


Figure 3.4. Conventional printed strip Yagi-Uda antenna configuration.

A design that serves as a wide-band Yagi-Uda antenna or as a multi-band printed dipole

was realized by using parasitic elements in proximity to a dipole that is parallel to a truncated ground plane [60, 61]. Recently, the effects of all the antenna parameters for such antennas on the frequency response were investigated in [62].

Table 3.2 shows the optimized parameters for the 'standard' Yagi-Uda antenna shown in Fig. 3.4 to resonate in the DTV frequency band between 475 MHz and 794 MHz [63].

Table 3.2. Optimized parameters of the conventional Yagi-Uda antenna to cover the DTV frequency bandwidth. The dimensions are explained in Fig. 3.4.

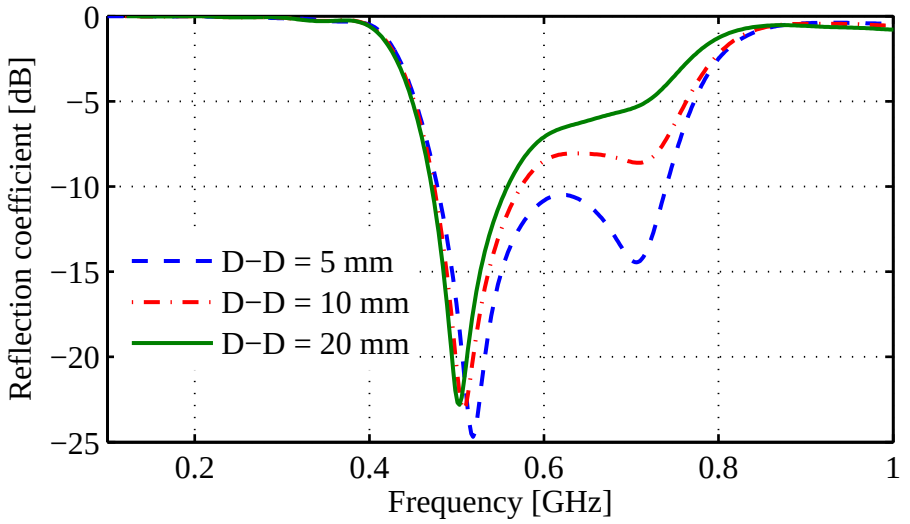
Parameter	mm
F-L	235
F-W	15
R-L	290
R-W	30
D-D, D-L	see Sec. 3.3.1
R-D	see Sec. 3.3.2

The Yagi-Uda antenna parameters including the length of the director (D-L), the distance between the feed and the director (D-D), and the distance between the feed and the reflector (R-D) are optimized to obtain the widest bandwidth possible that can cover the DTV broadcasting frequency bandwidth (475 MHz - 794 MHz). The feed length F-D is set to 235 mm so that the Yagi-Uda antenna resonates in the required frequency band (475 MHz - 794 MHz). The reflector length R-L is set to 290 mm as indicated in Table 3.2.

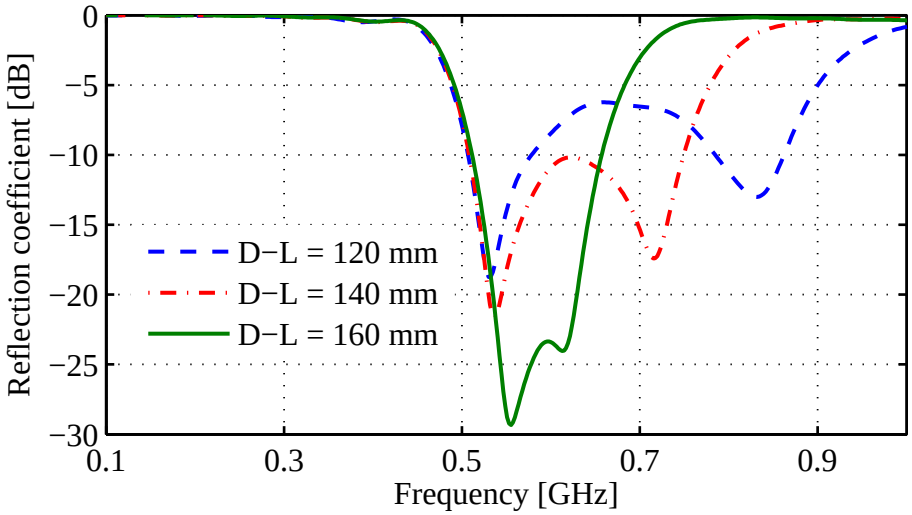
3.3.1 Optimization of the Distance Between Feed and Director (D-D) and the Director Length (D-L)

In this section the distance between feed and director (D-D) and the director length (D-L) are optimized to broaden the frequency response of the conventional printed Yagi-Uda antenna. The distance between the reflector and the feed is fixed to 110 mm and will be optimized in the following section.

Figure 3.5(a) shows the simulated reflection coefficient as a function of frequency for different distances between the feed and the director. It is clear from the figure that, when the director is set close to the feed, a second resonance is generated at a higher frequency which enhances the reflection coefficient bandwidth. This is due to the increased mutual coupling between the feed and the director. Actually the director acts as a resonator of length $\lambda/4$ when it is close to the driven element.



(a)



(b)

Figure 3.5. (a) Reflection coefficient versus frequency for different D-D, D-L = 140 mm, R-D =110 mm. And (b) reflection coefficient versus frequency for different D-L, R-D = 110 mm, D-D = 5 mm.

To confirm this observation, the distance between the director and the driven element is set to D-D= 5 mm, and the director length is swept between 120 mm and 160 mm. As expected, see Fig. 3.5(b), the reflection coefficient is characterised by two resonances. The low-frequency resonance is due to the feeding length (F-L) designed for 530 MHz, the high-frequency resonance is due to the length of director (D-L). The latter resonance shifts toward the lower frequency resonance when the length of the director is increased. This behaviour makes it possible to bring the two resonances together, which results in an improved reflection coefficient over an extended frequency band.

3.3.2 Optimization of the Distance Between Feed and Reflector (R-D)

Based on the parameters in Table. 3.2, the optimization of the distance between the feed and the reflector (R-D) is presented in this section.
Figure 3.6 shows the reflection coefficient as a function of frequency for different distances between the feed and the reflector.

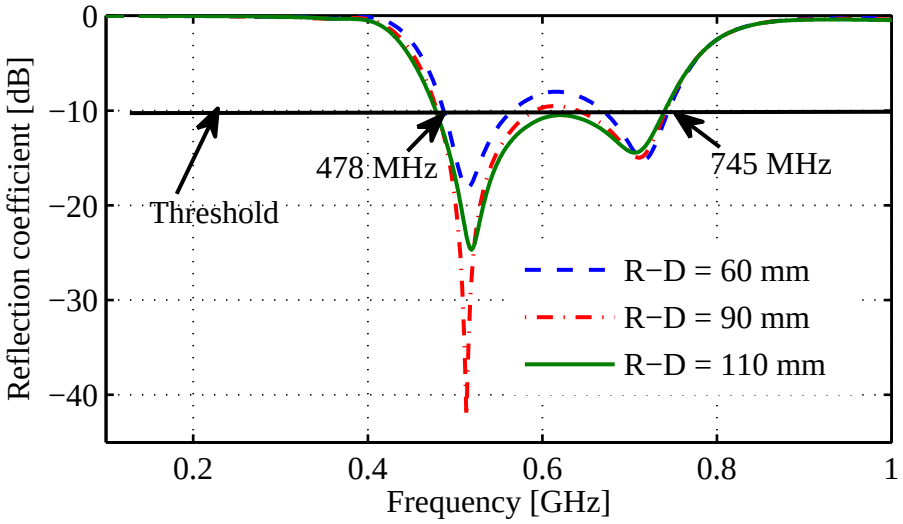


Figure 3.6. Simulated results of the reflection coefficient as a function of frequency for different distances between the reflector and the feed. D-L = 140 mm, D-D = 5 mm.

It is clear from the simulation results that the operating antenna bandwidth is not strongly affected by the distance between the feed and the reflector. However, an increased reflection coefficient exceeding the threshold of -10 dB (dashed curve, R-D = 60 mm), is observed when the reflector is placed nearby the feed. When the distance between the feed and the director is enlarged, the absolute value of the reflection coefficient decreases as illustrated by the solid curve for R-D = 110 mm.

After all the Yagi-Uda antenna parameters have been optimized, it is shown by the solid curve in Fig. 3.6 that the maximum achievable bandwidth is 267 MHz (478 MHz - 745MHz) which does not meet the full bandwidth requirements (475 - 794 MHz).

3.4 Modified Yagi-Uda antenna

To further broaden the frequency band and at the same time preserve the radiation characteristics, a novel broadband modified Yagi-Uda antenna is introduced in this sec-

tion.

As was shown in Sec. 3.2, the step-shaped-fed strip dipole is characterized by two resonant modes, the first one being due to the strip length and the second one due to the total effective length of the antenna comprising its two asymmetric radiator arms 1 and 2 [55] (see Fig. 3.3). We have also seen that by carefully placing the parasitic elements in a 3-elements Yagi-Uda antenna, a dual resonance may be obtained. In this section the driven element in Fig. 3.4 will be replaced by the step-shaped-fed strip dipole antenna shown in Fig. 3.2(b). The resulting configuration is shown in Fig. 3.7.

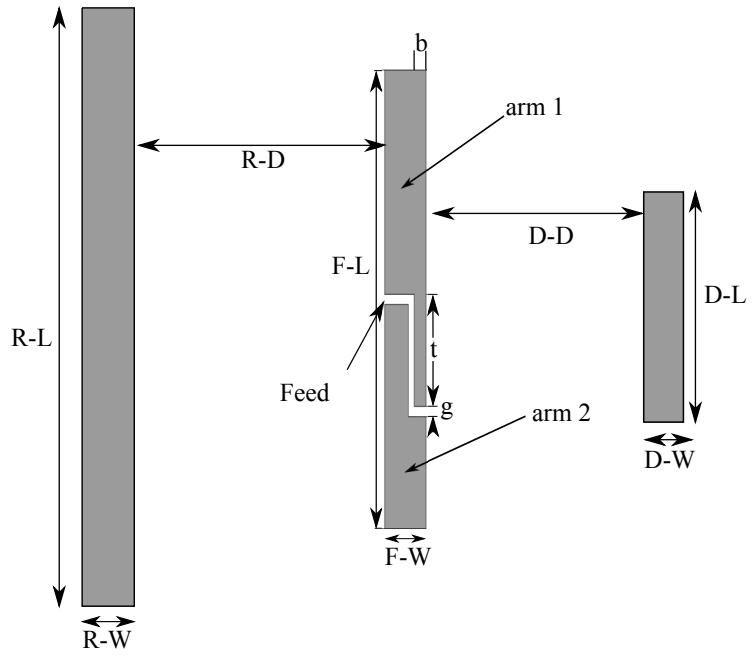


Figure 3.7. Modified Yagi-Uda antenna with a step-shaped-fed driven element.

It is the expected that this modified configuration will be characterised by three resonant modes, one due to the feed length (F-L), a second one due to the director length (D-L) and a third one due to the total effective length of the feed compromising its two asymmetric radiators arms 1 and 2 (0.5F-L + t). The main advantage of this configuration will be that it offers three additional parameters b, t and g (see Fig. 3.2(b) and Fig. 3.7) for resonance frequency tuning. By optimizing these additional parameters the frequency bandwidth will be widened to cover the complete DTV band (475 MHz - 794 MHz).

To validate that the modified Yagi-Uda antenna is characterized by three different resonant modes, that can be tuned independently, CST Microwave Studio is used to simulate the reflection coefficient versus frequency. The previously designed Yagi-Uda antenna is used as a starting point. The dimensions are summarized in Table 3.3. Only

the parameters, D-L, b and t will be sequentially optimized.

Table 3.3. Optimized parameters of the modified Yagi-Uda antenna to cover the DTV frequency band.

Parameter	mm
F-L	235
F-W	15
R-L	290
R-W	30
R-D	110
D-D	5
g	1
D-L	see Sec. 3.4.1
b	see Sec. 3.4.2
t	see Sec. 3.4.3

3.4.1 Modified-Yagi: Three Resonant Modes

In this section the director length is optimized to demonstrate that the modified Yagi-Uda is characterized by three resonant modes. Figure 3.8 shows the reflection coefficient as a function of frequency for different director lengths (D-L). The parameters shown in Table 3.3 are used for simulations.

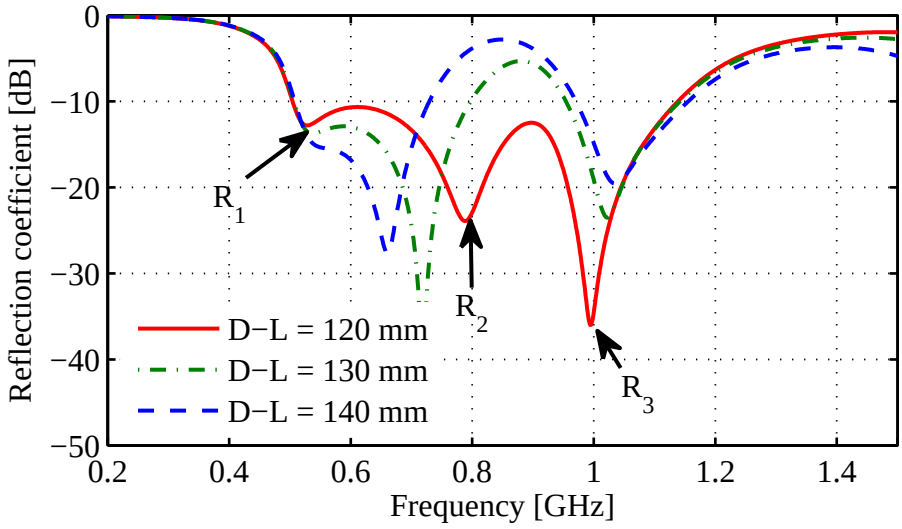


Figure 3.8. Optimization of the novel broad band Yagi-Uda antenna parameter D-L. Simulated results of the reflection coefficient as a function of frequency. b = 1 mm, t = 35 mm, g = 1 mm, R-L = 235 mm.

As expected, the figure indicates that the modified Yagi-Uda antenna is characterised by three resonance frequencies, R_1, R_2 and R_3. The first resonance frequency band around R_1 is due to the driven element length F-L. The second resonance R_2 is due to the length of the director D-L and shifts toward the first resonance when the length of the director increases. The third resonance R_3 is due to the effective length of the feed that is controlled by the arm width b and the step length t.
Possessing three resonant modes that can be tuned independently, the modified Yagi-Uda antenna is potentially a broad-band antenna if these three resonant frequencies are close enough. If the three resonant frequencies are separated, the modified Yagi-Uda antenna operates as a triple-band antenna. If two of the three resonant frequencies are close enough the antenna operates as a dual-band antenna.

3.4.2 Optimization of the Arm Width b

In this section the effect of the parameter b is discussed. The arm width is swept between 3 mm and 7 mm since the strip width F-W (see Fig.3.7) is fixed to 15 mm. Figure 3.9 shows the reflection coefficient as a function of frequency for the different arm widths b.

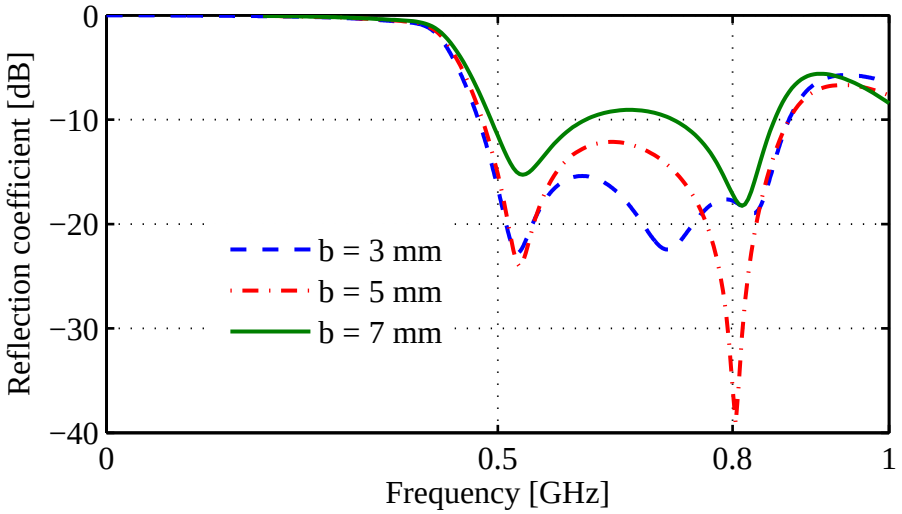


Figure 3.9. Optimization of the novel broad band Yagi-Uda antenna parameter b. Simulated results of the reflection coefficient as a function of frequency for different values of b. $g = 1$ mm, $t = 45$ mm, D-L = 130 mm.

It is observed that b has a strong effect on the reflection coefficients. It is shown in the figure that a smaller width b will lead to a lower reflection coefficient. When b increases, the absolute value of the reflection coefficient increases to surpass the -10dB threshold for $b = 7$ mm. Thus, we decide to fix b at 3 mm.

3.4.3 Optimization of the Step Length t

In this section, the step length t is sequentially optimized so that the frequency bandwidth of the modified Yagi-Uda antenna covers the required DTV frequency band. Figure 3.10 shows the simulated reflection coefficient as a function of frequency for different step lengths t. As expected, the parameter t has a low impact on the lower resonance position (R_1). The higher resonance mode (R_3) is clearly affected and is shifted towards the lower frequency resonance R_2, when the step length t increases.

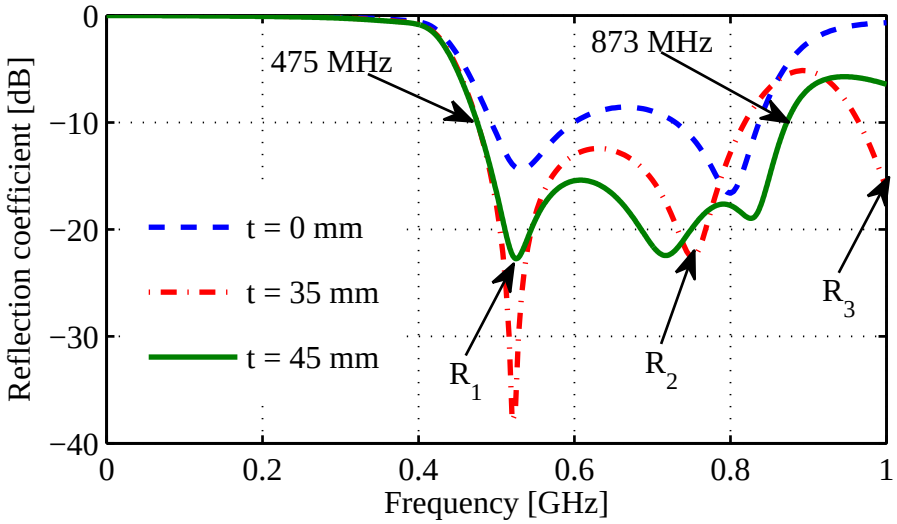


Figure 3.10. Simulated results of the reflection coefficient as a function of frequency for different step length values t. $g = 0.5$ mm, $b = 3$ mm, D-L = 130 mm.

As a result, a wide bandwidth is achieved when t is set to 45 mm. When compared to a standard centre fed Yagi-Uda antenna (dashed curve, $t = 0$ mm) which is characterized by only two resonances, the reflection coefficient of the novel antenna does not exceed the -15 dB level between 595 and 850 MHz, while the reflection coefficient of the standard Yagi-Uda antenna reaches -8.7 dB at a frequency of 680 MHz.
The sequentially optimized parameters for covering the DTV bandwidth are as follows: gap width $g = 1$mm, step length $t = 45$ mm, arm width $b = 3$mm and director length D-L = 130 mm. For these parameters, the antenna impedance bandwidth reaches 398 MHz (475 MHz - 873 MHz) or about 61 % in 2.0:1.0 VSWR centered at 650 MHz, compared to 50 % in 2.5:1.0 VSWR for the wide-band strip dipole antenna in [55].
The gain and the front-to-back ratio as a function of frequency of the optimized antenna are shown in Fig. 3.11.
The novel antenna shows a gain above 4.3 dBi in the operating bandwidth. The front-to-back ratio is less than 9.6 dB to reach the DTV broadcasting range. Above 810 MHz the front-to-back ratio degrades, and reaches 5.2 dB at a frequency of 875 MHz.

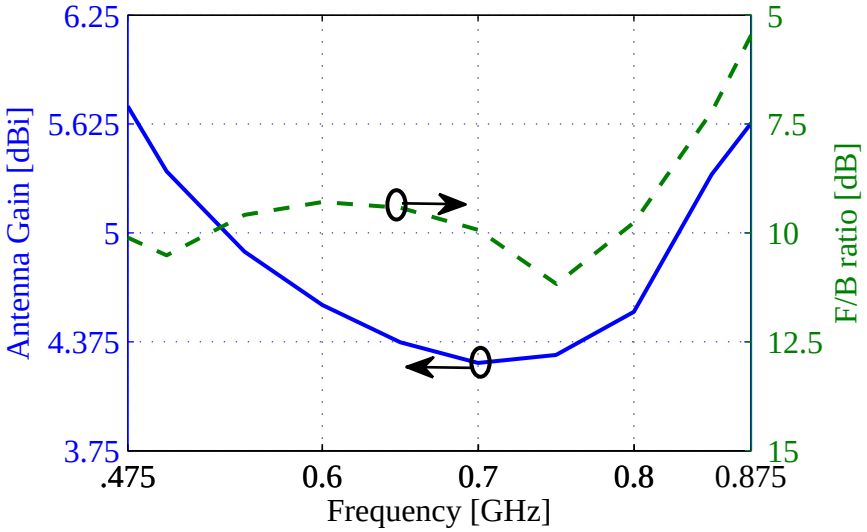


Figure 3.11. Simulated antenna gain and front-to-back ratio as a function of frequency for the novel broad-band Yagi-Uda antenna antenna, $b = 3$ mm, $t = 45$ mm, D-L = 130 mm. Other parameters are listed in Table.3.3.

The antenna gain for the new configuration has almost doubled with respect to the step-shaped-fed strip dipole antenna investigated in Sec. 3.2, which makes the newly proposed antenna well suitable for RF energy harvesting from DTV broadcast signals. In Table. 3.4 the comparison between the performances of a standard Yagi-Uda antenna and the modified Yagi-Uda antenna are presented.

Table 3.4. Comparison between a standard Yagi-Uda antenna and the modified broadband Yagi-Uda antenna.

	Standard Yagi-Uda (See Fig. 3.5(a)-solid curve)	Modified Yagi-Uda [This work]
Gain	5.7 dBi	4.2 dBi
F/B ration	10 dBi	5.2 dB
Bandwidth	80 MHz	398 MHz
Antenna size	290 mm x 295 mm	290 mm x 190 mm

Compared to a standard printed Yagi-Uda antenna (See Sec. 3.3) where the bandwidth around the resonance frequency is limited to 80 MHz (See Fig. 3.5(a)-solid curve), the novel antenna operating bandwidth is increased by a factor of four. The antenna gain over the wide band is only reduced by an average of 1.5 dBi. The front-to-back ratio

for a standard Yagi-Uda antenna is 5 dB lower, which makes the standard Yagi-Uda antenna more directive. However, the novel antenna (290 mm x 190 mm) is 35 % smaller when compared to a standard Yagi-Uda antenna (290 mm x 295 mm). The antenna gain of the novel broadband Yagi-Uda antenna can be improved at the expense of the antenna bandwidth and size.

3.4.4 Fabricated Antenna

Based on the previous results, an antenna is fabricated as shown in Fig. 3.12.

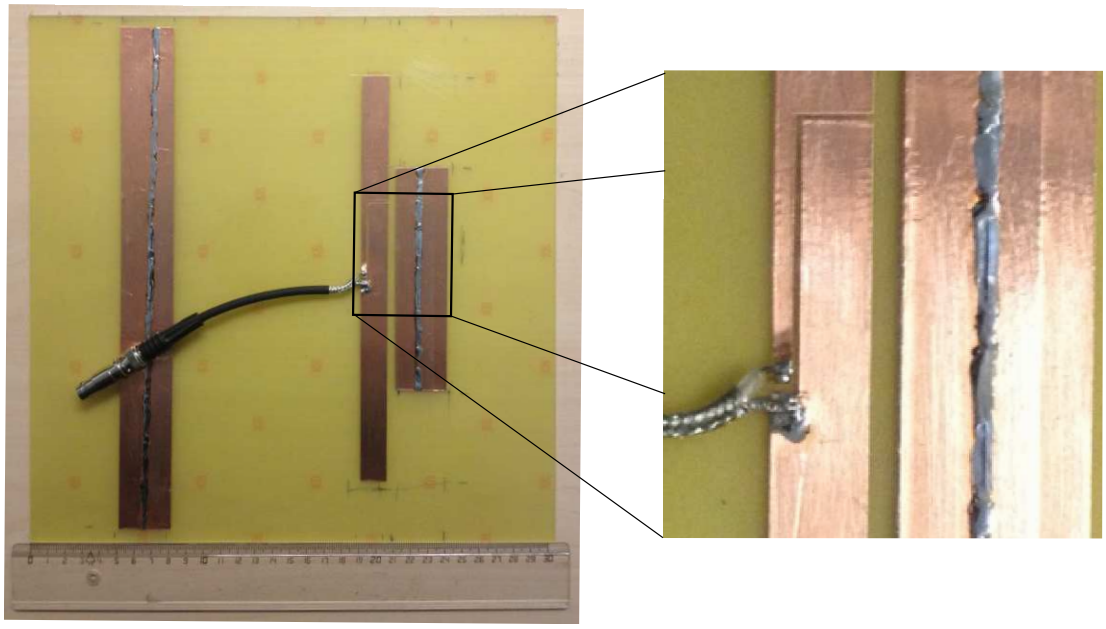


Figure 3.12. Fabricated antenna on a 1.6 mm thick FR4 substrate.

The antenna is printed on a 1.6 mm thick FR4 substrate. The reflection coefficient of the fabricated antenna is measured between 10 MHz and 1 GHz with the aid of the PNA-X Network Analyser [45]. The measured reflection coefficient as a function of frequency is shown in Fig. 3.13. It is clear form the figure the measured results of the reflection coefficients are lower than -10 dB for the frequency band between 475 MHz and 950 MHz which validates the broadband characteristics for the novel antenna.

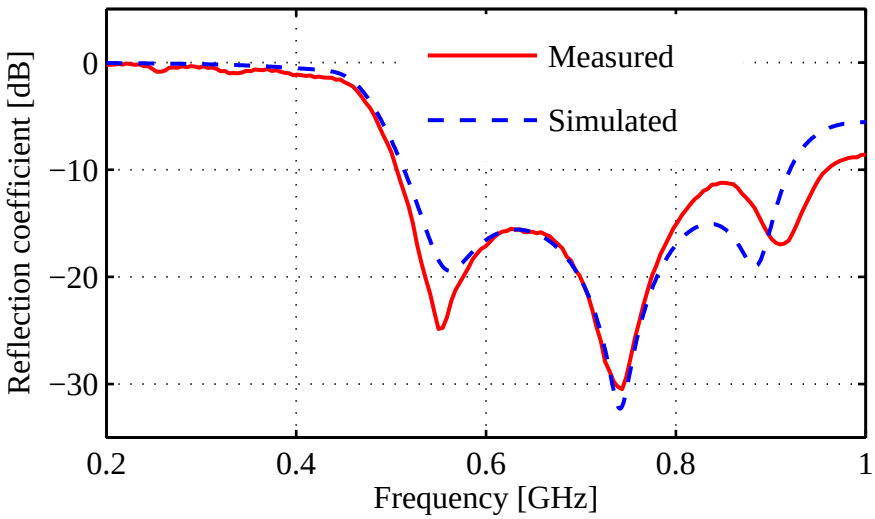


Figure 3.13. Simulated and measured reflection coefficient as a function of frequency.

Radiation diagrams in the E-plane of the fabricated prototype are measured at three different frequencies, 470 MHz, 650 MHz and 800 MHz. The E-plane and H-plane are indicated in Fig. 3.14.

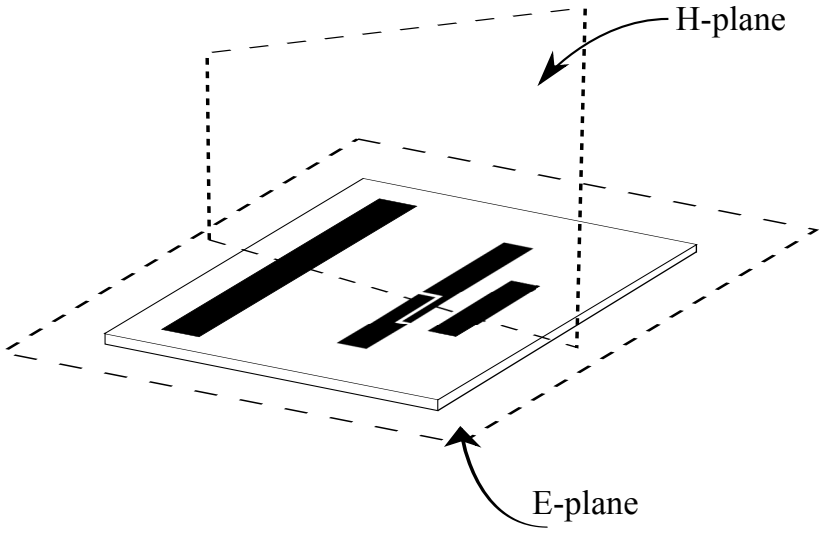
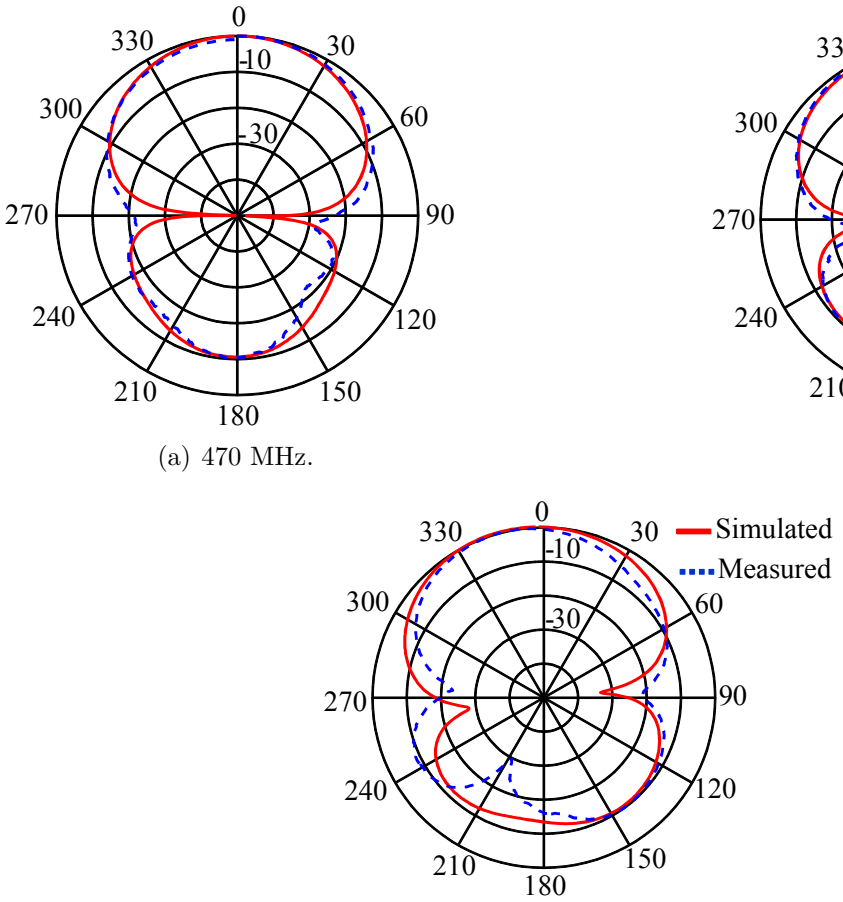


Figure 3.14. E-plane and H-plane.

Figure 3.15 shows the simulated and the measured radiation diagrams in the E-plane at the three different frequencies. The solid curve represents the simulated radiation diagrams while the dashed curve represents the measured ones. It is clearly indicated in the figure that the simulated radiation diagrams are validated by measurements with less than 0.5 dB difference in the direction of the main beam between $\theta = 330°$ and $\theta =$

30° at 470 MHz and 650 MHz. At 800 MHz, the difference increases to 2.5 dB at 30°, but note that both in simulations and measurements the main beam has been tilted to the left.



(a) 470 MHz.

(b) 650 MHz.

(c) 800 MHz.

Figure 3.15. Simulated and measured normalized radiation diagrams in the E-plane of the broad-band Yagi-Uda antenna.

3.5 Triple-Band Modified Yagi-Uda Antenna

The exponential increase in the usage of wireless communication devices, especially mobile phones, increases the presence of electromagnetic radiation in the environment. The dominant communication systems are as follows: GSM (890-910 MHz), DCS/PCS/UMTS (1800-2100 MHz) and WiFi (2.446 - 2.454 GHz). RF energy harvesting from these transmitting sources has gained a lot of attention recently. A multi-band RF power harvesting system was introduced in [64]. In this section, the triple-band antenna suggested in [64] to have the possibility to harvest RF power at three different frequencies (900 MHz, 1800 MHz and 2.45 GHz), is presented.

Using the same design procedures as in the previous section, the antenna parameters can be optimized to produce a triple-band or a dual-band antenna. This section introduces a triple-band modified Yagi-Uda antenna. The following section introduces a dual-band modified Yagi-Uda antenna. The optimized parameters (see Fig. 3.7) for the triple-band antenna prototype are shown in the first two columns of Table 3.5.

Table 3.5. Optimized parameters for the triple-band and for the dual-band prototypes. The antenna is fabricated on a 1.6 mm thick FR4 substrate.

Parameter	Triple-band (mm)	Dual-band (mm)
D-L	42	38
F-L	130	119
F-W	10	10
R-L	170	160
R-W	5	5
R-D	20	20
D-D	6	5
D-W	8	5
g	1.2	0.5
b	1.35	2
t	13	11

The triple-band antenna (see column 1 and 2 in Table 3.5) is designed, simulated and manufactured on a 1.6 mm thick FR4 substrate. The simulated (solid curve) and the measured (dashed curve) reflection coefficients as a function of frequency are shown in Fig. 3.16.

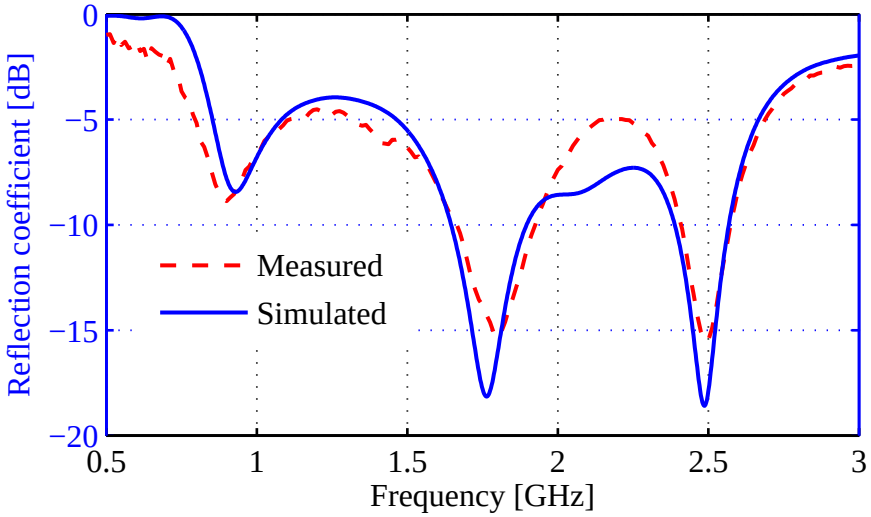
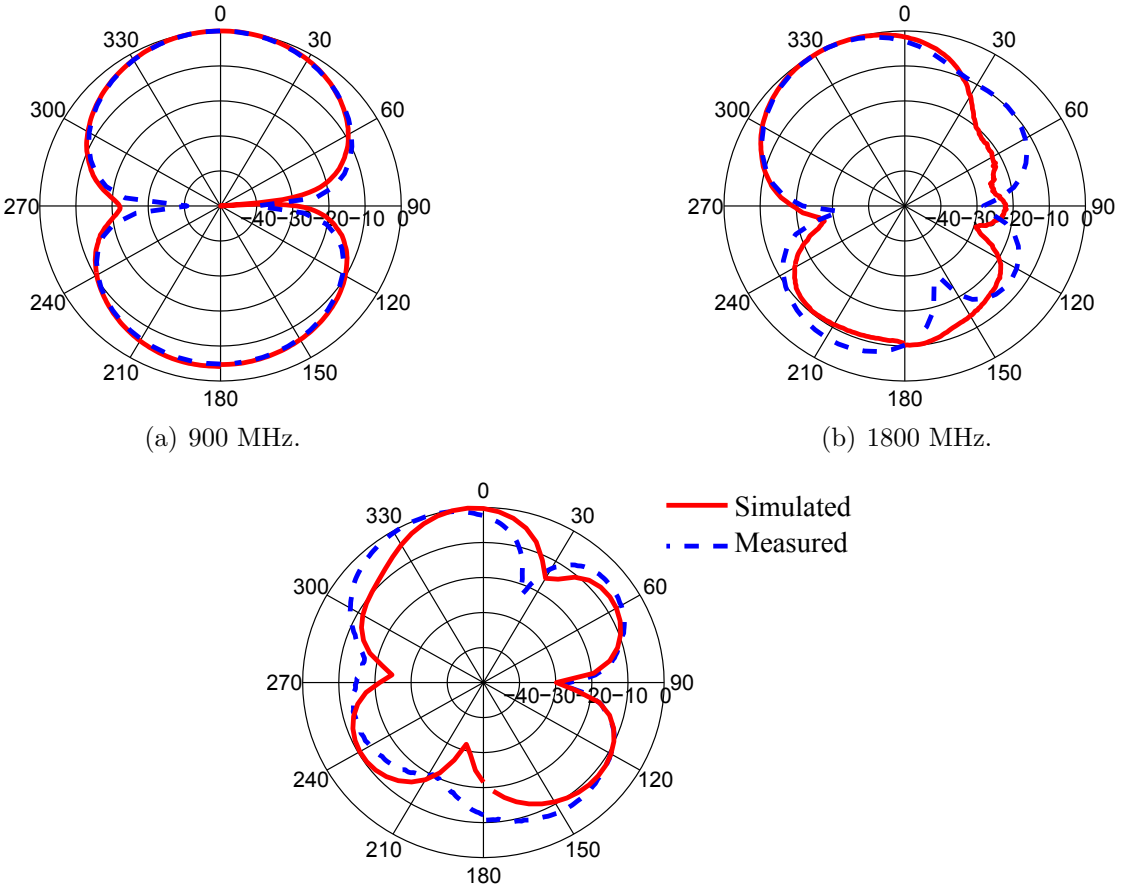


Figure 3.16. Measured and simulated reflection coefficients versus frequency of the triple-band Yagi-Uda antenna

It is observed in the figure that the measured reflection coefficients agree well with the simulated ones. The differences between the simulated and the measured results are less than 1 dB at 900 MHz, the difference increase to less than 3 dB at 1800 MHz and 2.45 GHz. The measured reflection coefficient is -8.6 dB, -16 dB and -13.25 dB at 900 MHz, 1800 MHz and 2.45 GHz respectively. This validates the triple-band characteristics for the antenna.
The normalized power radiation diagrams in the E-plane (see Fig. 3.14) have been simulated and measured at the three desired frequencies and are shown in Fig. 3.17.



(a) 900 MHz.

(b) 1800 MHz.

(c) 2.45 GHz.

Figure 3.17. Simulated and measured normalized power radiation diagrams in the E-plane of the triple-band modified Yagi-Uda antenna.

Figures 3.17(a), 3.17(b) and 3.17(c) show the simulated and the measured normalized power radiation diagrams at 900 MHz, 1800 MHz and at 2.45 GHz respectively. It is shown that the simulated and the measured results agree well with less than 0.5 dB difference in the direction of the main beam between $\theta = 330°$ and $\theta = 30°$ at 900 MHz. It should be mentioned that the direction of the maximum propagation is shifted 30

degrees at 1800 MHz and at 2.45 GHz. This shift is due to the asymmetry in the driven element for the modified Yagi-Uda antenna (see Fig. 3.7). In the direction of the shifted main beam at 1800 MHz, between $\theta = 300°$ and $\theta = 0°$, the difference between the simulated radiation diagram and the measured one is 1 dB at 1800 MHz.
At 2.45 GHz, in the direction of the main beam $\theta = 330°$, the difference increases to 3 dB. Here the radiation pattern is more complex, showing multiple lobes explaining the higher differences.

3.6 Dual-Band Modified Yagi-Uda Antenna

To demonstrate the reconfigurability of the novel antenna geometry, a third prototype is presented in this section. The antenna here is designed to be a dual-band antenna resonating at 900 MHz and at 2.45 GHz. The optimized parameters for the novel dual-band antenna are listed in Table 3.5 (column 1 and 3).
Figure 3.18 shows the simulated (solid curve) and the measured (dashed-dotted curve) reflection coefficient versus frequency as well as the simulated antenna gain (dashed curve) as a function of frequency.

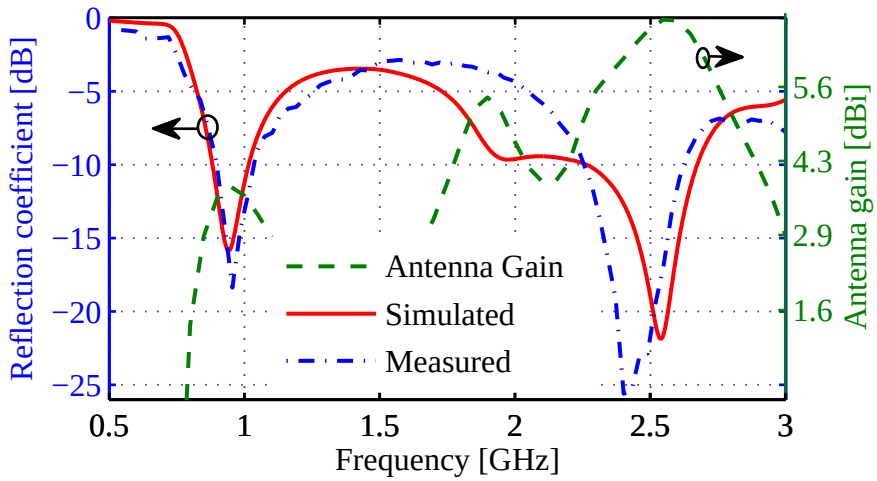


Figure 3.18. Simulated and measured results of the reflection coefficient and antenna gain as a function of frequency for the dual band antenna.

It is shown in the figure that a frequency shift of 100 MHz is present between the simulated and the measured reflection coefficients at the second resonance. This shift is due to fabrication tolerances due to the narrow gap width g of 0.5 mm. The simulated antenna gain reaches 3.2 dBi at 900 MHz and 6.45 dBi at 2.45 GHz.
The radiation diagrams in the E-plane have been simulated and measured for the dual-

band modified Yagi-Uda antenna. Figures 3.19 and 3.20 show the simulated and the measured normalized power radiation diagrams in the E-plane at the frequencies 900 MHz and 2.45 GHz respectively. The differences between simulations and measurements are less than 1 dB in the direction of the main beam between $\theta = 330°$ and $\theta = 30°$ at 900 MHz and at 2.45 GHz.

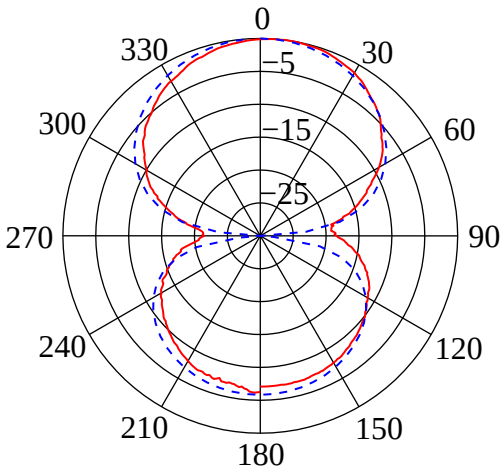


Figure 3.19. Simulated (dashed curve) and measured (solid curve) normalized power radiation diagrams in the E-plane of the dual-band modified Yagi-Uda antenna at 900 MHz.

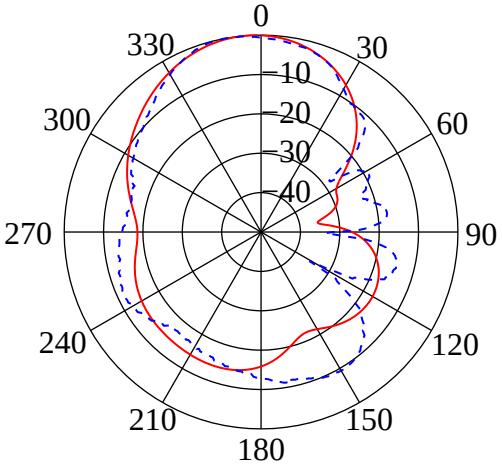


Figure 3.20. Simulated (dashed curve) and measured (solid curve) normalized power radiation diagrams in the E-plane of the dual-band modified Yagi-Uda antenna at 2.45 GHz.

3.7 RF Power Transport at 868 MHz

In this section, the receiving part of the RF power transport system is discussed. Design and measurement results will be described. The DC output voltage and the power conversion efficiency are measured for different values of the load resistance and for different input power levels.

3.7.1 Lumped-Element Matching Network

As is apparent from Fig. 3.1, a rectifier is needed to transform the RF input power into DC power. For a maximum power transfer between the antenna and the rectifier, an impedance matching network is designed, simulated, manufactured, and measured. Commercially available, discrete Schottky diodes HSMS-2852 [43] are used for power rectification.

Figure 3.21 shows a voltage doubler configuration. C_1 and C_2 are set to 100 pF. This value of C_1 and C_2 is not critical for the voltage doubler functioning. The capacitors need to function as 'DC-blockers'. R_L is set to 10 kΩ. The main reason behind choosing $R_L =$ 10 kΩ is that the measured power conversion efficiency peaks around 10 kΩ. Moreover, this value is around the optimum input impedance of the power management circuit [65] that will follow the rectifier part (see Fig. 3.1). $Z_{rectifier}$ is the input impedance of the voltage doubler.

The diode is a nonlinear device and its impedance changes as a function of frequency and as a function of input power level. A fixed input power level and a fixed operating frequency are chosen. Based on the results from previous chapters, we choose $P_{in} = -10$ dBm and the operating frequency $f = 868$ MHz (see Sec. 2.7). A lumped-element matching network is designed to match the input impedance of the rectifier to the input impedance of the antenna (50 Ω).

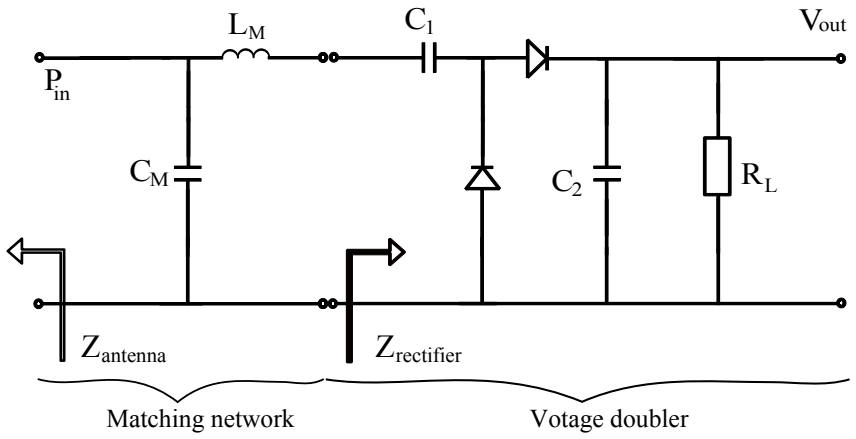


Figure 3.21. Voltage doubler configuration used to rectify the RF power harvested.

Agilent ADS [44] harmonic balance is used for impedance simulations. Figure 3.22 shows the Printed Circuit Board (PCB) footprint layout of the voltage doubler and the lumped-elements matching network. The components used to realize the circuit are summarized in Table 3.6. The circuit is printed on a 1.6 mm thick FR4 substrate.

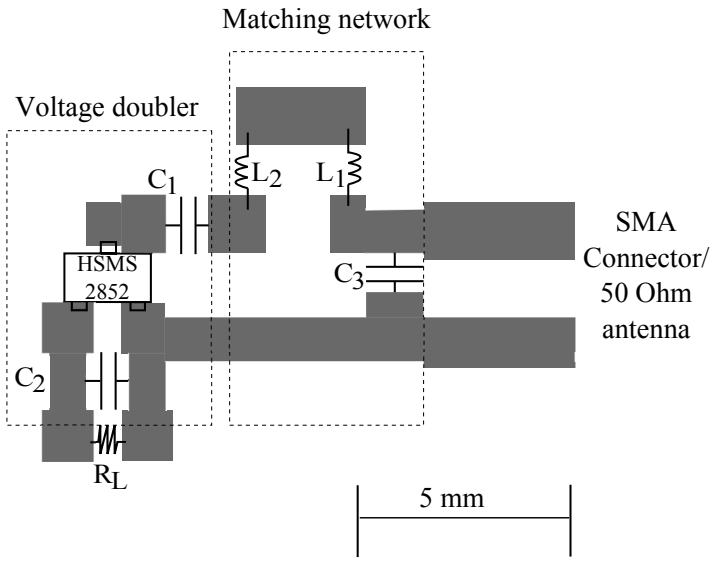


Figure 3.22. Layout of the PCB footprint of the voltage doubler and the matching network.

Table 3.6. Used lumped elements components.

Reference	Value	Manufacturer Ref.
C_3	0.7 pF	AVX SQCFVA 0R7BAT1A\500
C_2, C_1	100 pF	AVX 08052U101J
L_1, L_2	16 nH	JOHANSON TECHNOLOGY L- 15W16NJV4E
R_L	Variable resistor decade box	TENMA 72720
HSMS 2852	-	AVAGO TECHNOLOGIES HSMS-2852-TR1G

The reflection coefficient as a function of frequency for the realized network is measured for different input power levels, and the results together with the simulated ones are plotted in Fig. 3.23.

From the figures, it is observed that the simulated results agree reasonably well with the measured ones with a difference less than 3 dB between the simulated and the measured results. The designed matching networks match the impedance of the rectifier at 868 MHz for different power levels. At 868 MHz and using a 10 k Ω load the measured reflection coefficients are equal to -9.2 dB, -11.9 dB, -13.9 dB and -28.9 dB at -20 dBm, -15 dBm, -10 dBm and at -5 dBm, respectively. From the Figures, it is clear that the rectifier is matched to the antenna for the input power range -15 dBm $\leq P_{in} \leq -5$ dBm with a measured reflection coefficients less than -10 dB.

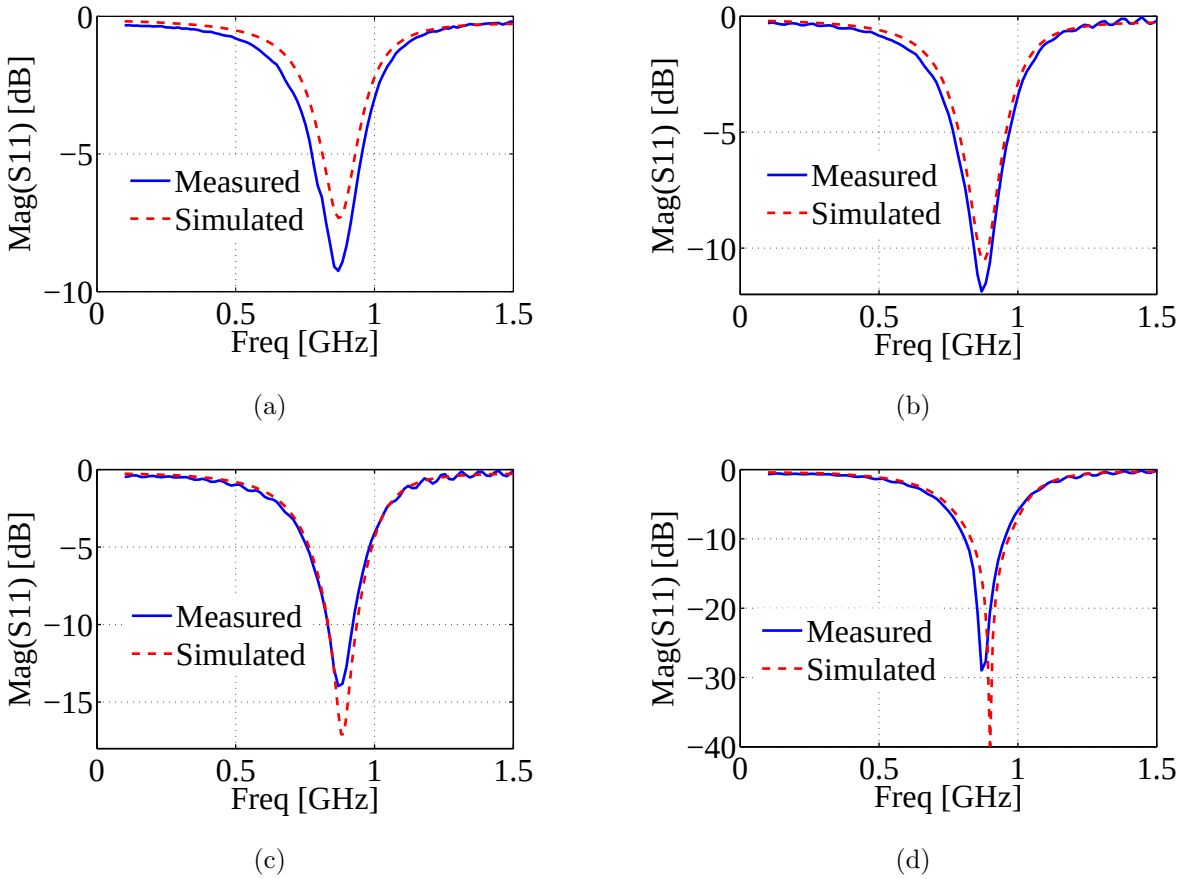


Figure 3.23. Reflection coefficients versus frequency for different input power levels. $R_L = 10$ kΩ. (a) $P_{in} = -20$ dBm . (b) $P_{in} = -15$ dBm. (c) $P_{in} = -10$ dBm . (d) $P_{in} = -5$ dBm.

3.7.2 RF Power Transport Measurement Results

For RF power transport, a system as shown in Fig. 3.1 is being used. In this section, we describe how the rectenna is used for wireless RF power transfer. The power

transfer measurement setup is shown in Fig. 3.24. The setup is calibrated by using two antennas with a known gain, G_r and G_t (see Fig. 3.24) at 868 MHz separated by 1 meter. With the choice of the antenna, we ensures that far-field conditions are satisfied for both antennas. The power density (S) at the receiving side is dictated by the effective isotropic radiated power, $EIRP = P_t G_t$, where P_t is the transmitted power and G_t is the transmit antenna gain, and by the distance R between the transmitting and the receiving antennas.

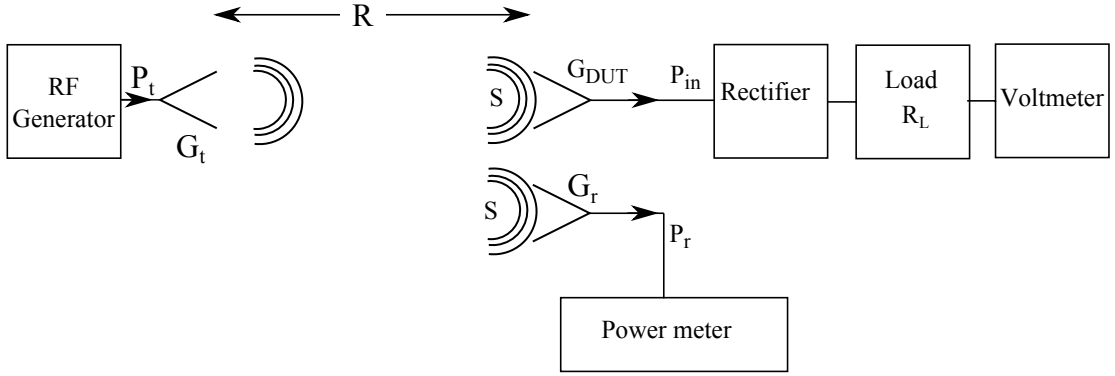


Figure 3.24. Power transfer measurement setup.

The power density is calculated by

$$S = \frac{EIRP}{4\pi R^2} = \frac{P_t G_t}{4\pi R^2}. \tag{3.1}$$

The antenna at the receiving side captures part of this power density with an effective aperture A_r. The received power P_r (see Fig. 3.24) is given by:

$$P_r = A_r S, \tag{3.2}$$

where A_r is the effective aperture of the receiving antenna in the direction of the transmitting antenna. According to [50], the effective aperture is related to the antenna gain by

$$A_r = G_r \frac{\lambda_0^2}{4\pi}, \tag{3.3}$$

where λ_0 is the wavelength of transmitted signal. Substituting Eq. (3.3) in Eq. (3.2) in Eq. (3.1) leads to the Friis transmission equation where the received power at the

terminals of the power meter is expressed as

$$P_r = P_t G_t G_r \left(\frac{\lambda_0}{4\pi R}\right)^2 = P_t G_t G_r \left(\frac{c}{4\pi f_0 R}\right)^2, \tag{3.4}$$

where c is the free-space speed of light and f_0 is the operating frequency of the transmitted signal.

A power meter is used to measure the received RF power P_r. The measured power P_r is then compared to the theoretically calculated value based on Eq. (3.4). The measured power appeared to be within ± 0.5 dB agreement with the theoretically calculated power, which ensures an accurate measurement setup. The rectenna is then placed for wireless power transport. The antenna gain of the rectenna is denoted by G_{DUT} as shown in Fig. 3.24. G_{DUT} is simulated with the aid of CST Microwave Studio. The input power level P_{in} at the clamps of the rectifier (see Fig. 3.24) is then calculated from

$$P_{in} = P_t G_t G_{DUT} \left(\frac{c}{4\pi f_0 R}\right)^2, \tag{3.5}$$

so that

$$P_{in} = \frac{P_r G_{DUT}}{G_r}. \tag{3.6}$$

The receiving rectenna system is connected to a variable load resistor R_L. The DC output voltage is measured with the aid of a voltmeter for different load resistances and for different input power levels.

The antenna of the receiving rectenna system shown in Fig. 3.24 consists of the modified Yagi-Uda antenna introduced in the previous section. The antenna is optimized to resonate at 868 MHz [37]. Then the antenna is connected to the rectifier through the impedance matching network and to a variable load resistor R_L. The DC output voltage is measured for different load resistances at three different input power levels -20 dBm, -10 dBm and -5 dBm calculated according to Eq. (3.6).

Figure 3.25 shows the measured output voltage versus load resistance at the three input power levels at an operating frequency f_0 of 868 MHz. The solid curve represents the output voltage as a function of R_L at an input power level of -20 dBm. The dashed curve and the dotted-dashed curve represent the measured output DC voltage at -10 dBm and -5 dBm respectively. The system output voltage reaches 0.17 V, 0.7 V and 1.3 V over a 10 kΩ load resistance at an input power level of -20 dBm, -10 dBm and -5 dBm, respectively.

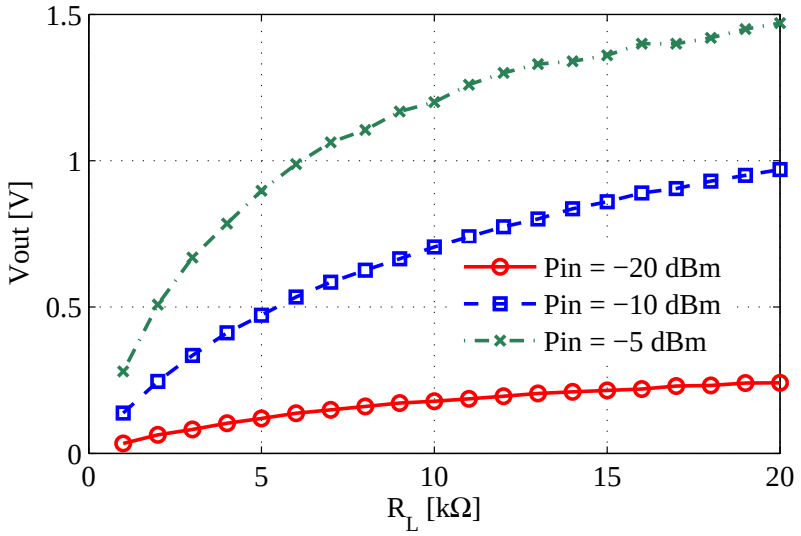


Figure 3.25. Measured DC output voltage versus load resistance R_L for different input power levels. Operating frequency f_0= 868 MHz.

The measured power conversion efficiency (PCE) is calculated using the following equation:

$$\mathrm{PCE}\,(\%) = \frac{P_{load}}{P_{in}} = \frac{V_{out}^2}{R_L}\frac{1}{P_{in}}. \tag{3.7}$$

The PCE is calculated for different input power levels from Eq. (3.7) and the results are plotted in Fig. 3.26.

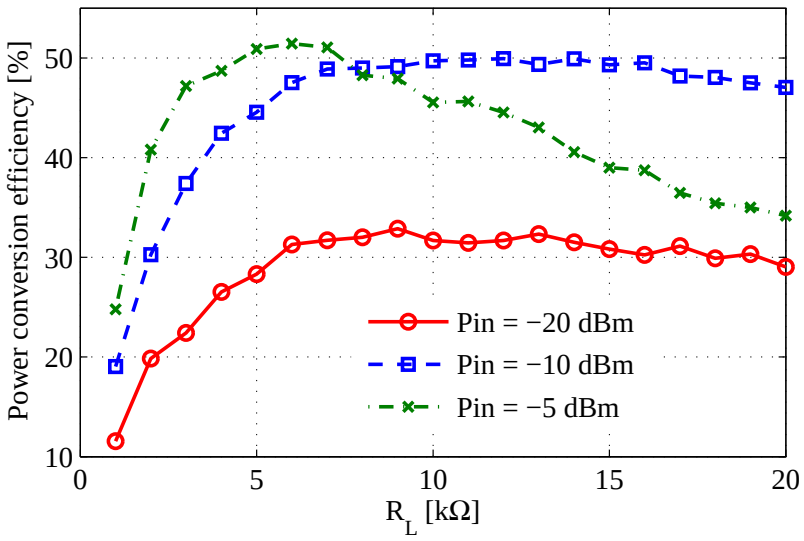


Figure 3.26. Measured power conversion efficiency (PCE) as a function of load resistance R_L for different input power levels. Operating frequency f_0 = 868 MHz.

The system PCE reaches 31.6% and 49.7 % over a 10 kΩ load resistance at an input power level of -20 dBm and -10 dBm, respectively. The PCE decreases to 45.5 % at an input power level of -5 dBm across the same load resistance. Table 3.7 compares the presented results in this section to the state-of-the-art reported results [19, 66–68] that use commercially available discrete rectifiers.

Table 3.7. Comparison of rectenna systems in literature that uses commertially available discrete rectifers

Reference	Pin [dBm]	Frequency [MHz]	Resistive Load kΩ	Discrete rectifier	Power Conversion Efficiency
[66]	-10	866.5	3	HSMS285C	24*%
[19]	-10 -20	1800	2.4	HSMS2850	38*% 15*%
[67]	-10	830	10^4	HSMS286Y	44%
[68]	-20	850 1850	-	SMS7630	14*% 13*%
This work	-10 -20	868	10	HSMS2852	49.7% 31.6%

∗calculated from figure

The authors in [66] and [19] use the same Schottky rectifier as was used in this section. The peak measured power conversion efficiency is 24 % across a 3 kΩ load in [66] and is 38% across a 2.4 kΩ load in [19] which is lower than the 49.7 % achieved by using the suggested matching network.

In [67], the authors use the commercial rectifier HSMS-286Y in combination with an impedance matching circuit and a resonant circuit. The maximum measured PCE at 830 MHz and at -10 dBm is 44 %. Compared to the results reported in Fig. 3.26, the system suggested in this chapter is 5 % more efficient. It is clearly indicated in Table 3.7 that the rectenna presented in this chapter shows a higher efficiency performance at -20 dBm and at -10 dBm.

3.8 Conclusion

A novel modified Yagi-Uda antenna is presented in this chapter. The main advantage of the presented antenna is that, by tuning its parameters, the antenna can be reconfigured to behave as either a broad-band, triple-band or dual-band antenna. Three prototypes are designed to validate the reconfigurability of the antenna. The antennas are simulated, manufactured and measured. The simulated results are in line with the

measured ones. Agilent ADS harmonic balance simulations are used to design a voltage doubler and a matching network to match the impedance of the rectifier to 50Ω.

The DC output voltage and the power conversion efficiency as a function of load resistance at three input power levels are measured for a frequency of 868 MHz. The system output voltage reaches 0.17 V, 0.7 V and 1.3 V over a 10 kΩ load resistance at an input power level of -20 dBm, -10 dBm and -5 dBm respectively. The system PCE reaches 31.6 % and 49.7 % over a 10 kΩ load resistance at an input power level of -20 dBm and -10 dBm, respectively.
The 50Ω rectenna system presented in this chapter exhibits an increased PCE of at least 5 % at -10 dBm and at least 10 % at -20 dBm, compared to the reported state-of-the-art results.

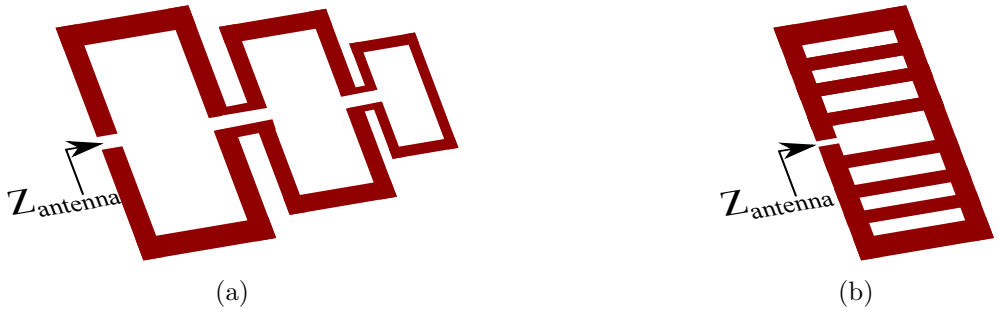
Chapter four

Compact and Conjugate-Matched Antenna Design

In this chapter we investigate different antenna topologies. These are the strip dipole, the folded strip dipole and a modified loop antenna. Novel, simple and accurate empirical expressions are introduced to calculate the input impedance of a strip dipole antenna. Next these expressions are employed to calculate the input impedance of a strip folded dipole antenna. A modified rectangular loop antenna with additional geometrical structures is introduced. The additional geometrical structures will make it possible to tune the antenna input impedance to the complex conjugate of that of a rectifier.

4.1 Introduction

To design an antenna with a specific input impedance, the geometry of the antenna should have enough parameters to be able to tune the input impedance. One such antenna is the strip-folded dipole array antenna shown in Fig. 4.1(a) [69]. This antenna is characterised by its broad bandwidth when the resonant frequencies of the individual elements are close together. If used for RFID applications and RF power harvesting, the strip folded-dipole array antenna has enough geometrical features to tune its input impedance. Another antenna that has ample geometrical features is the modified loop antenna shown in Fig. 4.1(b). Horizontal inner strips are added to the loop to tune the antenna input impedance. The main advantage of this modified loop antenna is its compact size compared to the strip-folded-dipole-array antenna.



(a) (b)

Figure 4.1. (a) Strip folded-dipole array antenna configuration and (b) the suggested modified loop antenna configuration.

The core of a strip folded-dipole array antenna is the folded dipole antenna. It is worthwhile therefore to put some effort into calculating the characteristics of the strip folded dipole antenna with a high degree of accuracy, without resorting to full-wave methods. The core of a folded dipole antenna is a dipole antenna. In this chapter, new empirical equations are introduced to calculate the input impedance of a strip dipole antenna. The proposed design equations are very accurate and can be easily implemented in standard computing tools, e.g. Matlab. The simulation results are validated by measurements. Next, updated analytical equations are presented to calculate the input impedance of an asymmetric strip folded dipole antenna. In addition, a novel and compact modified loop antenna configuration is introduced for which the antenna impedance can be tuned independently from its radiation efficiency. The impedance tuning ability eliminates the need for a matching network between the antenna and the rectifier which makes the rectenna system more compact, power-efficient, and cheaper to produce. The modified loop antenna will be used in Chapter 5 for RF power transport applications.

4.2 Strip Dipole Antenna

The basic element of the folded dipole antenna is the single dipole antenna [70]. A folded dipole antenna is commonly analyzed by recognizing a transmission line mode and a dipole antenna mode. The strip folded dipole antenna will be investigated in Sec. 4.3. In this section we present a strip dipole antenna analysis, using accurate and easy-to-use design equations to calculate the input impedance.

The input impedance of a dipole antenna can be calculated using full-wave analysis techniques, e.g. the method of moments (MoM) [71], and the finite integration technique (FIT) [72]. These methods are potentially very accurate for solving the Maxwell equations without a-priori approximations, but in general are time consuming. An alternative way is to use dedicated analytical-equation methods. Analytical equations may be derived by employing, among other methods, the induced EMF method, Hallén's

integral equation (HIE) method, and the King-Middleton second-order method [70]. While the resulting equations are easy-to-use, they appear to be insufficiently accurate to design antennas with a specific non-standard input impedance, especially when the radius of the cylindrical dipole antenna increases as shown in Appendix A. In this section, we present a simple but highly accurate analysis method that improves upon the known analytical expressions.

From a realization perspective, strip dipole antennas are more convenient than cylindrical wire dipole antennas. Strip dipole antennas may be implemented on foil, e.g. for RFID tags and even on-chip realizations are feasible.
A printed dipole antenna of width W can be analyzed by treating it as an equivalent cylindrical dipole antenna with a radius $r = \frac{W}{4}$ [73].

4.2.1 Equivalent Radius of a Strip Dipole Antenna

Cylindrical and strip dipole configurations are shown in Fig. 4.2(a).

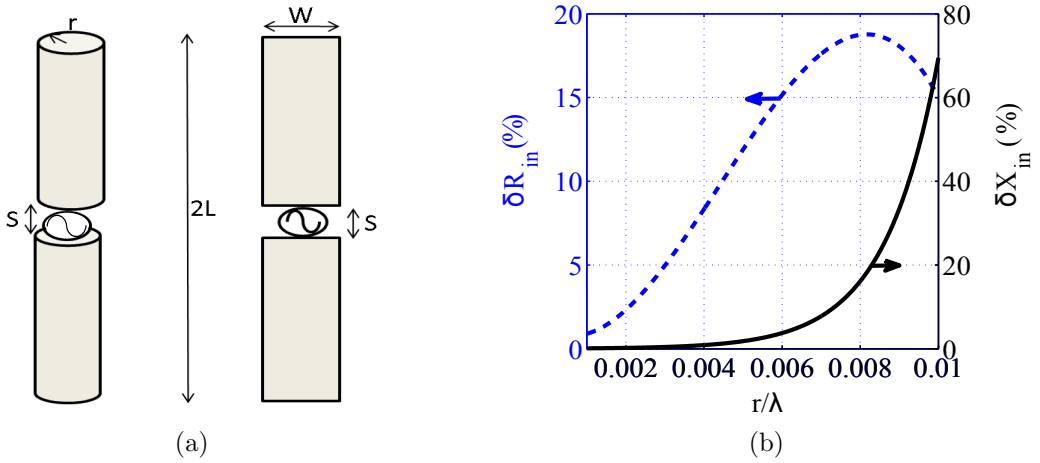


Figure 4.2. (a) Cylindrical and strip dipole configuration and (b) relative difference between the real and imaginary parts of the input impedance as a function of dipole radius over wavelength. L = 237 mm, S = 2 mm. $\lambda = 1$ m.

To investigate the accuracy of the equivalence $r = \frac{W}{4}$, the radius r of the cylindrical dipole is set to a quarter of the strip dipole width W, and CST Microwave Studio [57] is used to calculate the input impedance of the strip dipole and its equivalent cylindrical dipole. The feeding gap S is set to 0.002λ for the cylindrical dipole and for the strip dipole. This procedure is repeated for different dipole radii. Figure 4.2(b) shows the relative difference errors $\delta R_{in} = (R_{Cylindrical} - R_{Strip}) / R_{Cylindrical}$ (dashed curve) and $\delta X_{in} = (X_{Cylindrical} - X_{Strip}) / X_{Cylindrical}$ (solid curve) as a function of dipole

radius. It is shown in the figure that, for larger dipole radii, the deviation increases, which makes the quasi-static approach $\left(r = \frac{W}{4}\right)$ not accurate especially for larger dipole radii. Therefore, it should be used only for very thin cylindrical radii ($\frac{r}{\lambda} \leq 0.004$ for $\epsilon_{\delta_{R_{in},X_{in}}} \leq 10\,\%$). It should be mentioned that the frequency solver employed by CST Microwave Studio has been used for simulations which proved to be the most accurate compared to measurements.

4.2.2 Input Impedance of a Strip Dipole Antenna

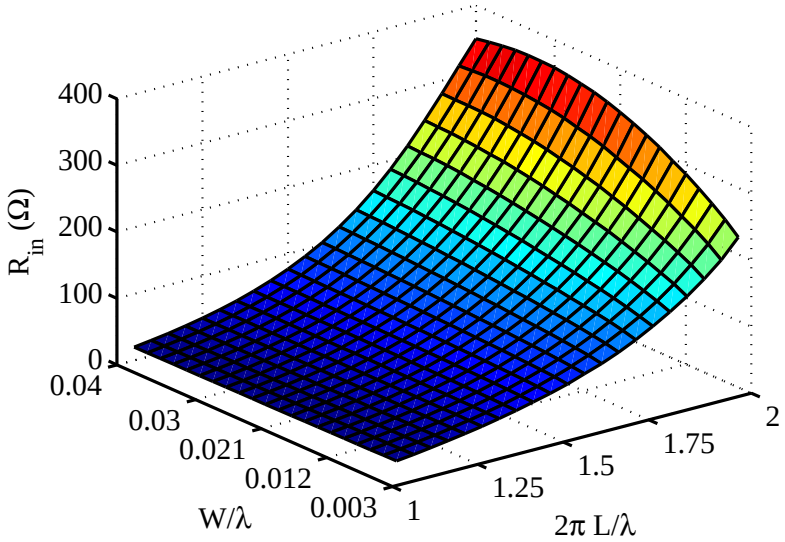
Since the quasi-static approach $\left(r = \frac{W}{4}\right)$ is not accurate enough to calculate the input impedance of a practical strip dipole, a new analytical solution to calculate the impedance of a strip dipole of length $2L$ and a width of W as shown in Fig. 4.2(a) is derived. The input impedance of the strip dipole antenna is calculated for different lengths and widths at a frequency of 300 MHz. Figure 4.3 shows the simulated results of the real (a) and imaginary (b) parts of the input impedance for $1.0 \leq \left(\frac{2\pi L}{\lambda}\right) \leq 2.0$ and $0.003 \leq \left(\frac{W}{\lambda}\right) \leq 0.04$ at an operating frequency of 300 MHz.
Different data fitting methods, using polynomial models, exponential models, Fourier series and power series have been investigated to best fit the results shown in Fig. 4.3. It has been found that the polynomial models match the simulated results with the lowest relative errors (less than 10 %). Matlab has been used to investigate the optimum fitting method. Since the strip dipole antenna has two parameters: the width W and the half-length length L, a two dimensional fitting expression is needed. Performing surface fitting on the full-wave calculated real an imaginary parts of the input impedances shown in Fig. 4.3, results in

$$R_{in}\left(\frac{L}{\lambda}, \frac{W}{\lambda}\right) = \sum_{m=0}^{5}\sum_{n=0}^{5} R_{mn}\left(\frac{L}{\lambda}\right)^{m}\left(\frac{W}{\lambda}\right)^{n}, \tag{4.1}$$

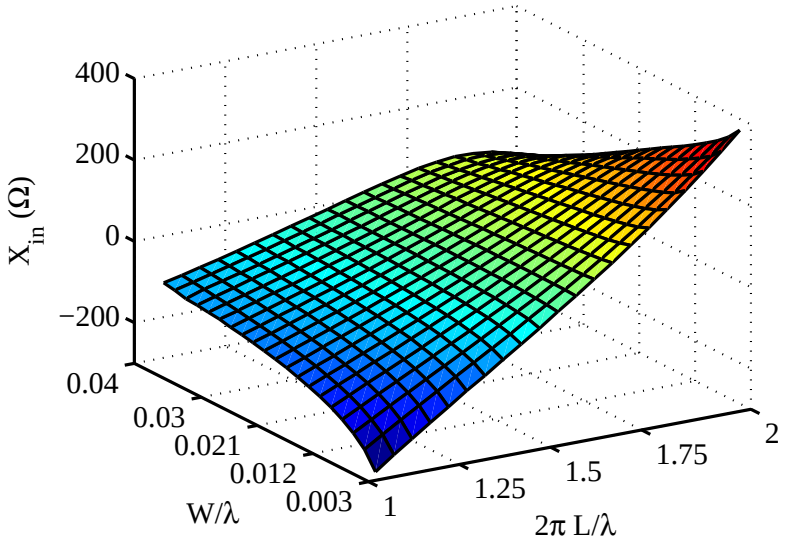
$$X_{in}\left(\frac{L}{\lambda}, \frac{W}{\lambda}\right) = \sum_{m=0}^{5}\sum_{n=0}^{5} X_{mn}\left(\frac{W}{\lambda}\right)^{m}\left(\frac{L}{\lambda}\right)^{n}, \tag{4.2}$$

where the coefficients R_{mn} and X_{mn} are listed in Table 4.1 and Table 4.2, respectively. The width of the feeding gap S is set to 0.002λ. The effect of the feeding gap is limited and will be discussed in detail in Sec. 4.2.3. The impedance equations are valid for $1.0 \leq \left(\frac{2\pi L}{\lambda}\right) \leq 2.0$ and $0.003 \leq \left(\frac{W}{\lambda}\right) \leq 0.04$, where they show a relative error less than 10 % for both real and imaginary parts. Compared to King-Middleton second-order fitting equations as described in Appendix A [70], which are valid only for $1.3 \leq \left(\frac{2\pi L}{\lambda}\right) \leq 1.7$ and $0.001 \leq \left(\frac{W}{\lambda}\right) \leq 0.01$, the novel expressions are valid for longer

and wider strip dipole antennas, as will be shown in Sec. 4.2.3. These equations can be used for a feeding gap width S in the range of 0.001λ to 0.005λ.



(a)



(b)

Figure 4.3. Real (a) and imaginary (b) parts of the input impedance of a strip dipole antenna as a function of antenna half-length (L) and antenna width (W) calculated by the Finite Integration Technique (FIT). $\lambda = 1$ m.

Table 4.1. R_{mn} coefficients used in Eq. (4.1) to calculate the input resistance of a strip dipole.

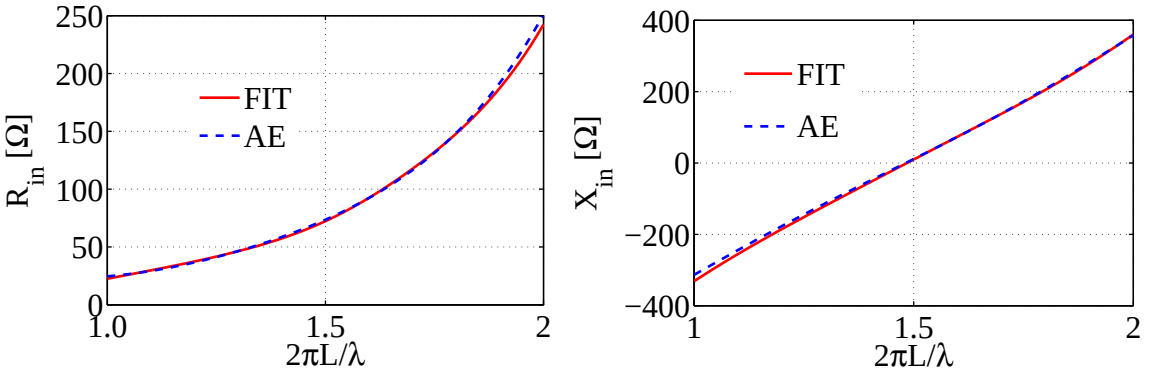
m\n	0	1	2	3	4	5
0	211.4	-3.456e4	7.96e5	-5.796e6	2.535e7	2.089e8
1	-1486	5.072e5	-9.993e6	4.49e7	-2.478e8	0
2	-1.544e4	-2.582e6	4.235e7	-5.274e7	0	0
3	2.047e5	4.834e6	-6.439e7	0	0	0
4	-7.511e5	-1.251e6	0	0	0	0
5	9.655e5	0	0	0	0	0

Table 4.2. X_{mn} coefficients used in Eq. (4.2) to calculate the input reactance of a strip dipole.

m\n	0	1	2	3	4	5
0	1022	-4.528e4	4.563e5	-2.015e6	4.277e6	-3.472e6
1	4.256e4	4.843e5	-6.868e6	2.478e7	-3.084e7	0
2	-4.147e6	2.13e7	-2.259e7	5.027e6	0	0
3	9.215e7	-3.885e8	2.088e8	0	0	0
4	-8.8e8	2.709e9	0	0	0	0
5	1.873e9	0	0	0	0	0

Figure 4.4 shows the real (a) and imaginary (b) parts of the input impedance of a strip dipole as a function of frequency calculated using the finite integration technique (FIT) and the analytical equations for strip dipoles (AE). The electrical length of the dipole is set to $\frac{2\pi L}{\lambda}$ =1.5 ($L = 11.93\,\text{cm}$), the width W is set to 0.003λ ($W = 1.5\,\text{mm}$). It is clear from Fig. 4.4(a) and Fig. 4.4(b) that the results obtained by the new empirical equations overlap the results obtained by the electromagnetic simulation software, which shows the accuracy of these fitting equations.

For this example, the relative difference error is calculated for both the real and imaginary parts. For the real part, the highest relative difference error occurs at a frequency of 800 MHz and it is equal to 8 %. For the imaginary part the highest relative difference error occurs at a frequency of 400 MHz and it is equal to 6 %. These equations have been verified for different strip widths and for different strip lengths as a function of frequency within the validity range. They show a relative error less than 10 % for both real and imaginary parts.



(a) (b)

Figure 4.4. Real (a) and imaginary (b) parts of the input impedance of a strip dipole antenna as a function of frequency calculated by the Finite Integration Technique (FIT) and by the novel Analytical equations (AE). $L = 11.93$ cm, $W = 1.5$ mm, $S = 1$ mm.

4.2.3 Feeding Gap S Dependence

The presented equations are developed for a fixed feeding width S of 0.002λ but may be valid over a range of S-values. The 0.002λ constraint may be too specific for general use. To investigate the effect of the feeding width, the FIT is used to calculate the reflection coefficients and the input impedance of two strip dipole antennas A1 and A2 with a length $L_1 = 225$ mm and a width $W_1 = 20$ mm for antenna A1. The length and width for antenna A2 are $L_2 = 115$ mm and $W_2 = 10$ mm, respectively.
Figures 4.5(a) and 4.5(b) show the reflection coefficients relative to 50Ω as a function of frequency for different feeding width values S for antennas A1 and A2, respectively. It is shown in the figures that the antenna resonance is shifted when the feeding width is changed. However, for practical values $(0.001\lambda - 0.005\lambda)$ the change in the reflection coefficient is negligible (see Figs. 4.5(a) and 4.5(b)). Consequently, for these practical values, the change in the impedance may be neglected.

To calculate the deviations between the input impedance calculated by the Analytical Equations (AE) and calculated by the Finite Integration Technique (FIT), the Figure of Merit (FoM) introduced in Sec. 1.6 is used. Using the FoM, the validity of the presented formulas for different feeding widths S are expressed in a single number having the characteristics of a power wave reflection coefficient. The input impedance of a strip dipole is calculated from the presented empirical expressions and by using the FIT for different feeding width values S. The FoM accounts for the mismatch due to feed gap errors in the analytical calculations.
Figure 4.6 shows the FoM for the analytical equations as a function of feeding gap width S referred to the outcome of the FIT analysis for antennas A1 (solid curve) and antenna

A2 (dashed curve). It is clear from the figure that the presented expressions are valid for a feeding width up to 0.02λ (Threshold$\leq -10\,\mathrm{dB}$), which validates the accuracy of the novel empirical expressions for the practical values of the feeding gap ($0.001\lambda - 0.005\lambda$) with a FoM $\leq$-20 dB.

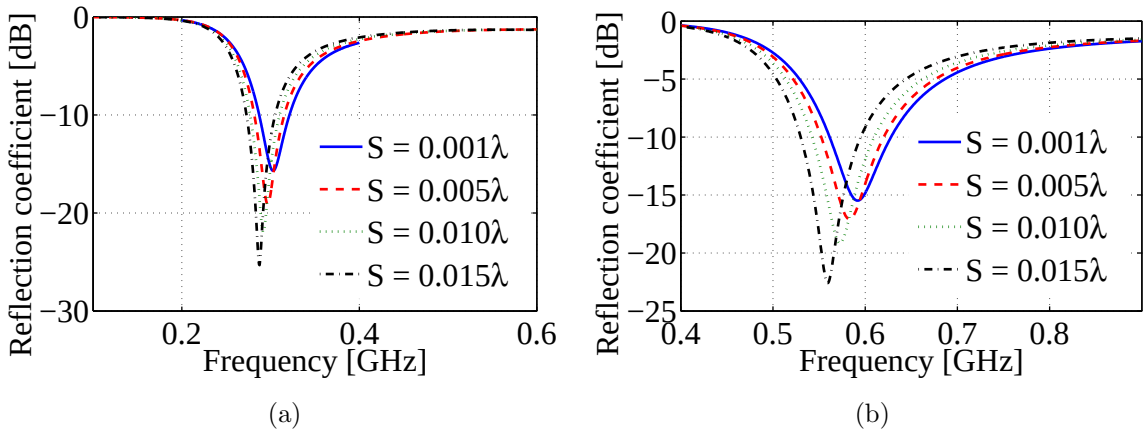


Figure 4.5. Reflection coefficients relative to 50Ω as a function of frequency for different feeding width S for antenna A1 (a) and antenna A2 (b). $L_1 = 225\,\mathrm{mm}$, $W_1 = 20\,\mathrm{mm}$, $\lambda_1 = 1\,\mathrm{m}$. $L_2 = 115\,\mathrm{mm}$, $W_2 = 10\,\mathrm{mm}$, $\lambda_2 = 0.5\,\mathrm{m}$

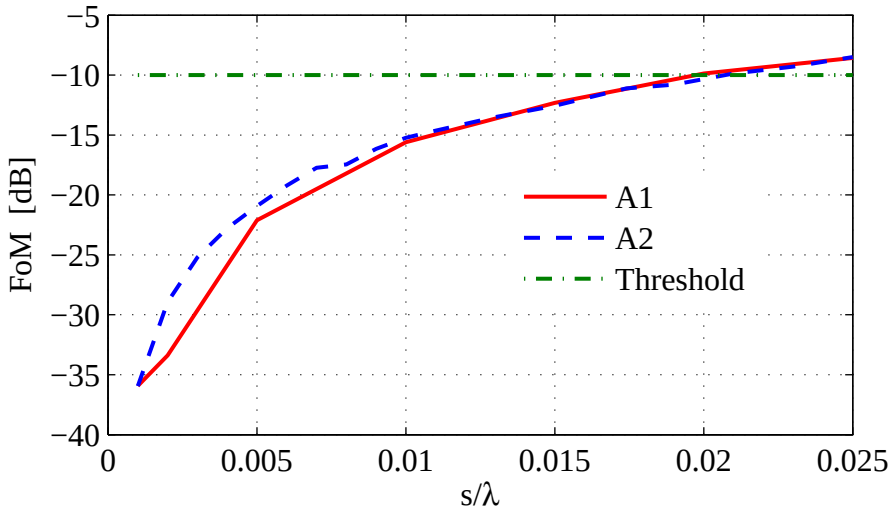


Figure 4.6. Figure of Merit (FoM) to calculate the deviations between the presented equations and the FIT for different feeding width for antenna A1 and antenna A2.

4.2.4 Fabricated Antenna

So far, we have used the results of CST Microwave Studio as a reference. To verify the real accuracy of the novel empirical equations, a strip dipole antenna is designed, simulated using AE and FIT, manufactured and tested. The antenna is designed to cover the GSM900 frequency bands: 890-915 MHz (uplink) and 935-960 MHz (downlink) [74]. The designed length and width of the strip dipole are $2L = 15\,\text{cm}$, $W = 1\,\text{cm}$. Figure 4.7 shows the simulated (AE & FIT) and the measured reflection coefficients as a function of frequency. It is clear from the full-wave analysis results and the measurement results that the novel empirical equations can predict the input impedance behavior and the resonance of the strip dipoles with less than 2 dB difference in the reflection coefficients in the specified GSM900 frequency bands.

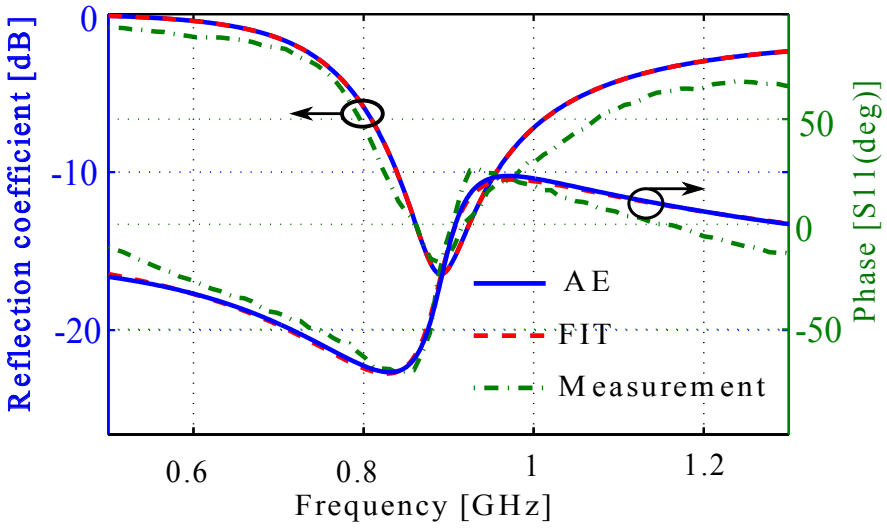


Figure 4.7. Simulated and measured reflection coefficients as a function of frequency.

4.3 Folded-Strip Dipole Antenna

The main limitation of the strip dipole antenna investigated in the previous section is its restricted number of geometrical features. This will make it hard to directly conjugate impedance match the antenna to a complex impedance. The impedance tuning ability of the strip dipole antenna depends only on the strip width W, the strip length $2L$ and - to a lesser extent - the feed gap width S (see Fig. 4.2(a)). Another antenna with more geometrical features that can be used for RF power harvesting applications is the strip folded dipole antenna [75] shown in Fig. 4.8(a). The strip folded dipole antenna may

be regarded as an evolution of the strip dipole antenna.

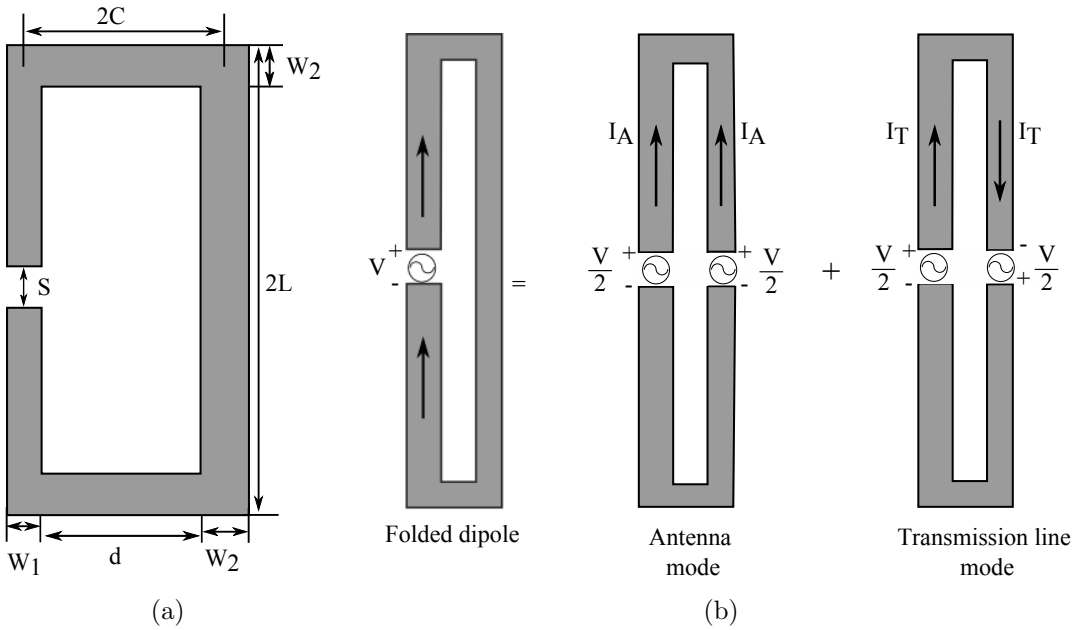


Figure 4.8. Strip folded dipole antenna configuration (a) and the decomposition of the current distribution into an antenna mode and a transmission line mode(b).

The main advantage of the strip folded dipole antenna over the strip dipole antenna is that it introduces two additional geometrical features, W_2, the width of the parallel, short-circuited dipole, and d, the inter-element distance between the CoPlanar Strips (CPS), (see Fig. 4.8(a).

In this section we will discuss an updated design procedure to calculate the input impedance of an asymmetric strip folded dipole antenna. Compared to the state of the art [76, 77], the design procedure in this section does not use the equivalence between a strip and a circular cross section wire to calculate the input impedance of a strip dipole. Instead it directly uses the empirical equations (Eq. (4.1) and Eq. (4.2)) for the strip dipole antenna introduced in the previous section.

The folded dipole antenna analysis by Thiele and Ekelman [75] will be followed. The currents in the folded dipole are decomposed into two fundamental modes: the transmission-line mode and the antenna mode as visualized in Fig. 4.8(b), where I_A is the antenna current flowing on the surface of the dipole and I_T is the transmission line current.

4.3.1 Input Impedance of a Strip Folded Dipole

Figure 4.8(a) shows the configuration of an asymmetric strip folded dipole antenna in free space. By decomposing the current into two modes, the input impedance of the strip folded dipole has become a function of three variables, the input impedance of the strip dipole, the input impedance of the coplanar strip transmission line stubs that may be recognized in the antenna configuration and the step-up impedance ratio [78,79] accounting for the difference in strip widths. The input impedance of the folded dipole antenna is then found to be [33]:

$$Z_{in} = \frac{2(1+q)^2 Z_D Z_T}{(1+q)^2 Z_D + 2Z_T}, \tag{4.3}$$

where Z_D is the input impedance of a strip dipole antenna with length $2L$ and width W_1. Z_T is the input impedance of the asymmetric coplanar strip transmission line stub of length L. The CPS line widths are W_1 and W_2, respectively, and $(1+q)^2$ is the step-up impedance ratio [75]. q will be specified in the next subsection.

4.3.2 Transmission Line Analytical Equations

The input impedance of a short circuited length of coplanar strip transmission line Z_T is calculated from the general transmission line theory as [80]:

$$Z_T = jZ_C \tan\left(\frac{\beta L}{2}\right), \tag{4.4}$$

where $Z_C = \left[\frac{120\pi}{\sqrt{\epsilon_r}} \frac{K(k)}{K'(k)}\right]$ is the characteristic impedance of the CPS transmission line [76]. $K(k)$ is the complete elliptic function of the first kind where $K'(k) = K(k')$ and $k'^2 = 1 - k^2$. The ratio of the complete elliptic functions $\frac{K(k)}{K'(k)}$ may be approximated by [81],

$$\frac{K(k)}{K(k')} \approx \begin{cases} \frac{1}{2\pi} \ln\left[2\frac{\sqrt{1+k}+\sqrt[4]{4k}}{\sqrt{1+k}-\sqrt[4]{4k}}\right] & \text{for } 1 \le \frac{K}{K'} \le \infty, \quad \frac{1}{\sqrt{2}} \le k \le 1 \\ \frac{2\pi}{\ln\left[2\left(\frac{\sqrt{1+k'}+\sqrt[4]{4k'}}{\sqrt{1+k'}-\sqrt[4]{4k'}}\right)\right]} & \text{for } 0 \le \frac{K}{K'} \le 1, \quad 0 \le k \le \frac{1}{\sqrt{2}} \end{cases} \tag{4.5}$$

The argument k is given by [78,79],

$$k = \frac{\left(\frac{d}{2}\right)\left(1 + e\left(\frac{d}{2} + W_1\right)\right)}{\left(\frac{d}{2}\right) + W_1 + e\left(\frac{d}{2}\right)^2}, \tag{4.6}$$

and

$$e = \left[\frac{W_1 W_2 + \frac{d}{2}(W_1 + W_2) - \sqrt{W_1 W_2 (d + W_1)(d + W_2)}}{\left(\frac{d}{2}\right)^2 (W_1 - W_2)} \right], \tag{4.7}$$

where W_1 is the width of the strip dipole, W_2 is the width of the parallel, short circuited dipole and d is the distance between the CPS strips as shown in Fig.4.8(a). The step-up impedance ratio $(1+q)^2$, is calculated from [78],

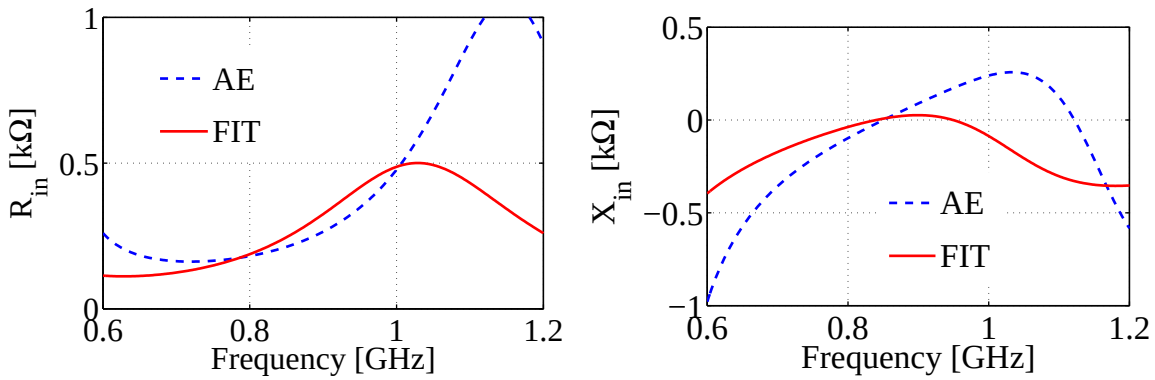
$$q = \left[\frac{\ln\left\{4C + 2\sqrt{(2C)^2 - \left(\frac{W_1}{2}\right)^2}\right\} - \ln\{W_1\}}{\ln\left\{4C + 2\sqrt{(2C)^2 - \left(\frac{W_2}{2}\right)^2}\right\} - \ln\{W_2\}} \right], \tag{4.8}$$

where $2C$ is the distance from the middle of the strip dipole to the middle of the parallel dipole as shown in Fig.4.8(a). ($C = \frac{d}{2} + \frac{W_1}{4} + \frac{W_2}{4}$). Now, we will follow the existing analysis for the strip folded dipole antenna and replace the dipole input impedance with the newly derived expressions.

4.3.3 Validation of the New Analytical Procedures

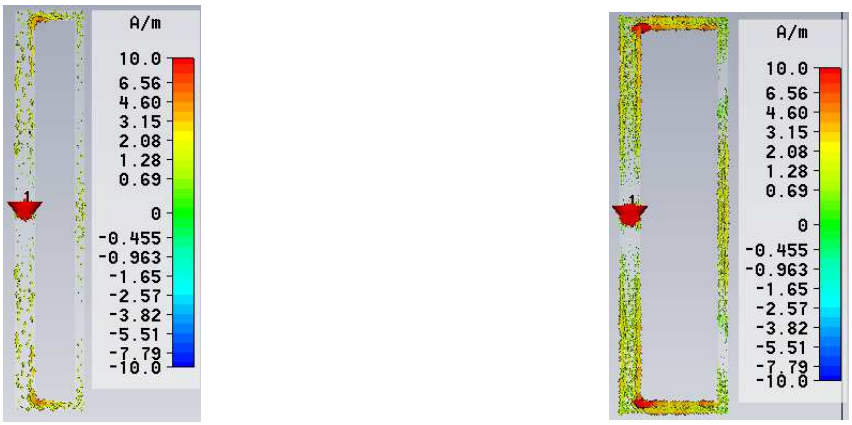
To verify the accuracy of the updated analytical expressions, the real and imaginary parts of the input impedance of a strip folded dipole are calculated using CST Microwave Studio (FIT). These results are compared to the results of the proposed analytical expressions (AE). Figure 4.9 shows the input impedance (real and imaginary parts) as a function of frequency for a strip folded dipole calculated by both the FIT and the AE. The strip widths are $W_1 = 8\,\text{mm}$, and $W_2 = 4\,\text{mm}$, the separation between strips is $d = 16\,\text{mm}$, the strip folded dipole length is $2L = 16\,\text{cm}$ and the feeding width is $S = 2\,\text{mm}$.

It is clear from the figure that the results obtained by the AE deviate from the results obtained by the FIT. Both curves (for the real and imaginary parts) show more or less the same behavior, but a frequency shift is visible as well as a change in the impedance level. To investigate possible sources of errors in the analytical expressions, the surface current distribution is simulated with the aid of CST Microwave Studio. Figure 4.10 shows the surface current distribution on a strip folded dipole antenna for two different separation distances, $d = 16\,\text{mm}$ and $d = 32\,\text{mm}$, at an operating frequency of 1.2 GHz.



(a) (b)

Figure 4.9. Real (a) and imaginary (b) parts of the input impedance versus frequency for an asymmetric strip folded dipole antenna calculated using the Finite Integration Technique (FIT) and the Analytical Equations (AE).$W_1 = 8\,\text{mm}$, $W_2 = 4\,\text{mm}$, $d = 16\,\text{mm}$, $2L = 16\,\text{cm}$.



(a) (b)

Figure 4.10. Surface current distribution on a strip folded dipole antenna for a separation between strips of (a) $d = 16\,\text{mm}$ and of (b) $d = 32\,\text{mm}$. $W_1 = 8\,\text{mm}$, $W_2 = 4\,\text{mm}$, $2L = 16\,\text{cm}$. Frequency $= 1.2\,\text{GHz}$

It is observed in Fig. 4.10 that a considerable amount of current flows along the short-horizontal strips. Moreover it is shown that when the separation distance is increased to $d = 32\,\text{mm}$ (Fig. 4.10(b)) the surface current density in these strips increases.
When the input impedance of the coplanar strip transmission line Z_T is calculated, it is assumed that the coplanar strip lines are connected with an infinitely thin horizontal strip, but in reality they are interconnected with a strip that has the same width as the transmission line W_2. Consequently, the deviations shown in Fig. 4.9 are assumed

to be due to the connecting strips of length d. In order to compensate for the distance between the CPS and the current flowing in the connecting strips, a correction to the transmission line length is required.
To derive the correction factor analytically, the input impedance of the antenna is calculated and compared to the results obtained by FIT. Then the length of the transmission line in the transmission line mode is increased until the resonance frequency of the AE coincides with that of the FIT. This procedure is repeated for different separation widths d.

The correction to the transmission line length L_T is calculated as:

$$L_T = 2L + \Phi(d)\,\lambda d, \tag{4.9}$$

where the correction factor $\Phi(d)$ for the separation width is found to be

$$\Phi(d) = \left(0.06243 d^{(-1.038)} + 3.071\right). \tag{4.10}$$

The results after using this correction factor are shown in Fig. 4.11. It is shown in the figure that when the correction factor is used, the proposed analytical equations can predict the impedance behavior of the strip folded dipole antenna within reasonable accuracy as will be explained in the following.

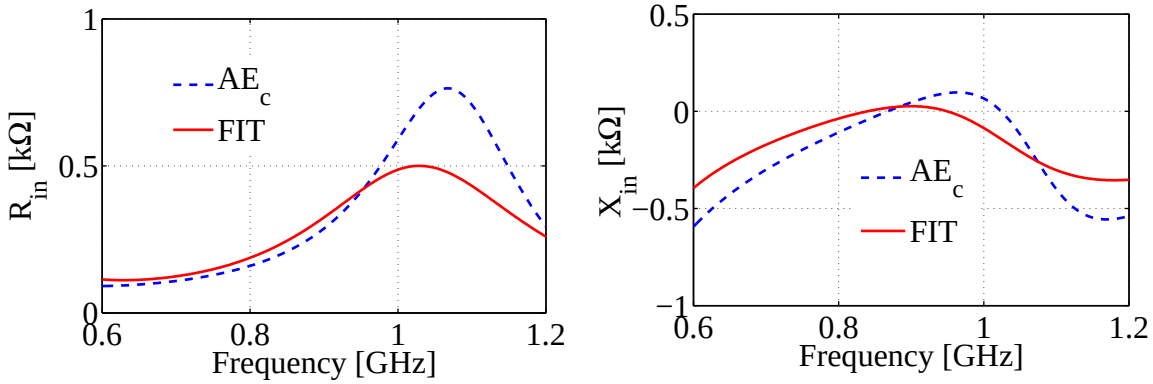


Figure 4.11. Real (a) and imaginary (b) parts of the input impedance versus frequency for the same asymmetric strip folded dipole antenna as analyzed in Fig. 4.9. The correction factor for the transmission line length is introduced in the analytical expressions.

To quantify the improvement of using the correction factor in terms of a single Figure of Merit, Eq.(1.2) is used to transform the deviations into power reflection coefficients. Using the suggested Figure of Merit, the deviations between the input impedance calculated by the FIT and by AE without a correction factor (Fig. 4.9) and when the

correction factor is employed (Fig. 4.11) are casted into a single plot. The higher the deviations between FIT and AE are, the higher the reflection coefficients (FoM) will be.

Figure 4.12 shows the Figure of Merit as a function of frequency for the two simulations. The dashed curve represents the FoM for the uncorrected Analytical Expressions (AE), while the solid curve represents the FoM for the corrected Analytical Expressions (AE_c). It is shown in the figure that, when the correction factor is used, the validity range of the corrected analytical expressions (0.75 GHz - 1.05 GHz) is extended by approximately 50 % compared to a validity range of the uncorrected analytical expressions (0.75 GHz - 0.95 GHz). The validity range is determined by a FoM not exceeding -10 dB. Two more validation examples, one for an antenna resonant at 300 MHz and one for an antenna resonant at 1.5 GHz, are presented in [82], and are shown in Appendix A.

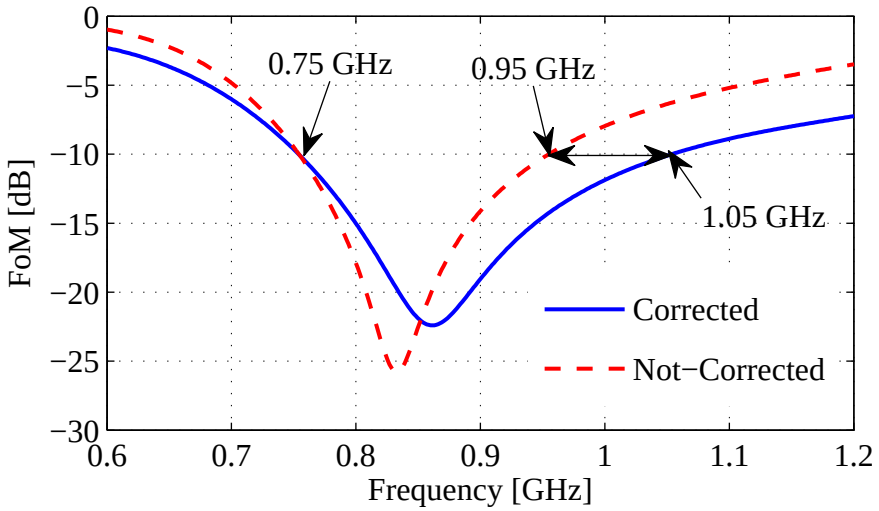


Figure 4.12. FoM as a function of frequency for the AE and corrected AE

The analytical expressions for the folded strip dipole antenna are valid within the range $1.3 \leq \left(\frac{2\pi L}{\lambda}\right) \leq 1.7$ ensuring a FoM $\leq$ -10 dB. The validity range of these expressions is also limited for d, being less than 1/20th of a wavelength.
Although we now have an antenna type with more possibilities to tune its input impedance than a strip dipole antenna, the tuning possibility may still be too limited for a practical use. Moreover, the size of the antenna (about half a wavelength long) might be too large for some applications, especially for Smart Building Integration (SBI). The following section therefore introduces a compact modified rectangular loop antenna.

4.4 Modified Rectangular Loop Antenna

In this section, a novel and compact antenna configuration is introduced to be conjugately matched to the impedance of the rectifier. The antenna is a modified rectangular loop antenna. The modification consists of adding additional arms to control its input impedance. The suggested antenna configuration is shown in Fig. 4.13 and consists of metal printed on a substrate. The radiation characteristics (radiation efficiency and radiation pattern) of the antenna are dictated by the outer antenna length and width L_1 and L_2, respectively, while adding additional arms to the antenna allows tuning its input impedance without affecting the radiation characteristics much. The width of each arm W_1 to W_n and the distance between the arms E_1 to E_n are fine tuned to obtain the exact desired input impedance.

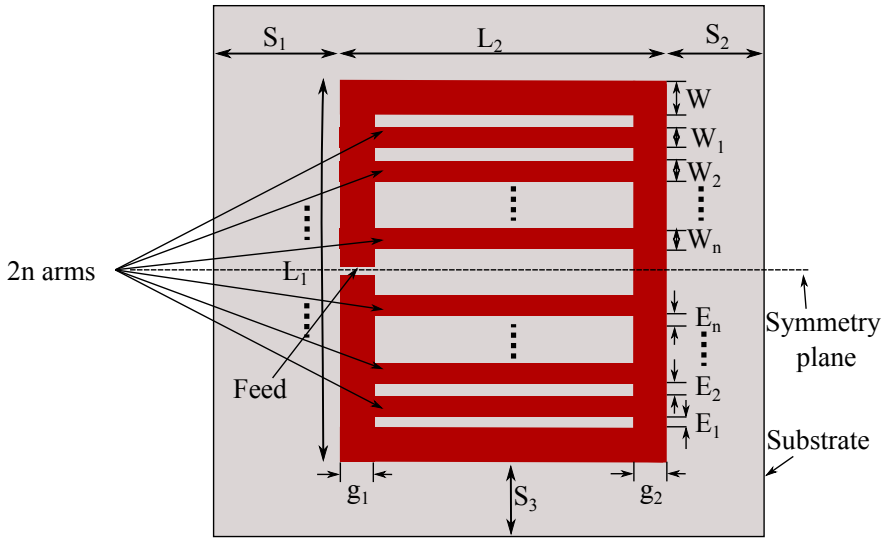


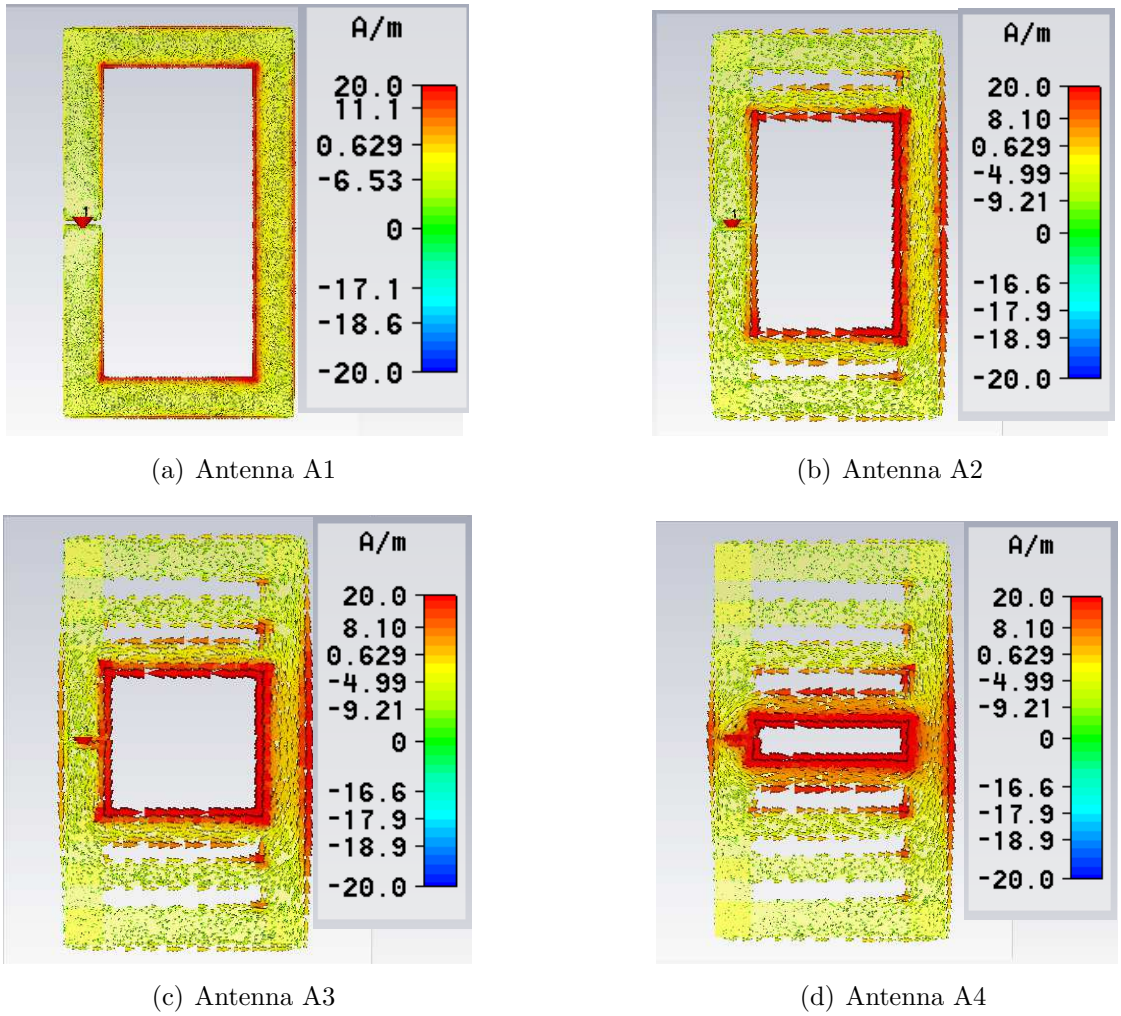
Figure 4.13. Compact antenna topology and dimensional parameters.

In order to understand the physical behavior of the suggested antenna, CST Microwave Studio is used to simulate the surface current density for four different antenna configurations that possess the same outer dimensions L_1 and L_2 (see Fig. 4.13). The parameters of the four simulated antennas denoted by A1, A2, A3 and A4 are listed in Table. 4.3.

Table 4.3. Antenna parameters used to calculate the input impedance and the radiation efficiency. All dimensions are in millimeters.

W_1, W_2, W_3, E_1, E_2	E_3	W, g_1, g_2	L_2	L_1	S_1	S_2	S_3
3	4	5	32	54	10	7	8

The first simulated antenna (A1) is a loop antenna without any additional arm see Fig. 4.14(a). The outer dimensions of the loop antenna are L_1 and L_2. The width of the strips are W, g_1 and g_2. The second simulated antenna (A2), with $n = 1$, has exactly the same outer dimensions as antenna A1 but with two additional arms added as shown in Fig. 4.14(b). The additional needed parameters to simulate antenna A2 are W_1 and E_1. In the same way the third antenna (A3) and the fourth antenna (A4) are designed and simulated. Antenna (A3) with $n = 2$, possesses the same outer characteristics as antenna A1 but has four additional arms while antenna A4 ($n = 3$) has 6 additional arms compared to antenna A1.



(a) Antenna A1

(b) Antenna A2

(c) Antenna A3

(d) Antenna A4

Figure 4.14. Surface current density on the suggested compact loop antenna for different numbers of additional arms. Frequency = 868 MHz.

Figure 4.14(a) shows the surface current distribution for antenna A1. It is clear from the figure that the surface current density is distributed along the loop. When two

arms are added to antenna A1 to form antenna A2, the maximum current density is concentrated in the inner loop as shown in Fig. 4.14(b). Similarly, when more internal arms are added to the loop, the maximum current density is drawn by the inner loops as shown in Fig. 4.14(c) and Fig. 4.14(d). Using the suggested configuration, the antenna input impedance and its radiation efficiency can be tuned separately. Since the surface current density is concentrated in the inner loop, consequently the inner loop dictates the input impedance of the antenna, while the outer loop dimensions L_1 and L_2 determine the antenna radiation efficiency. The arms between the inner loop and the outer loop are used to fine tune the antenna input impedance.

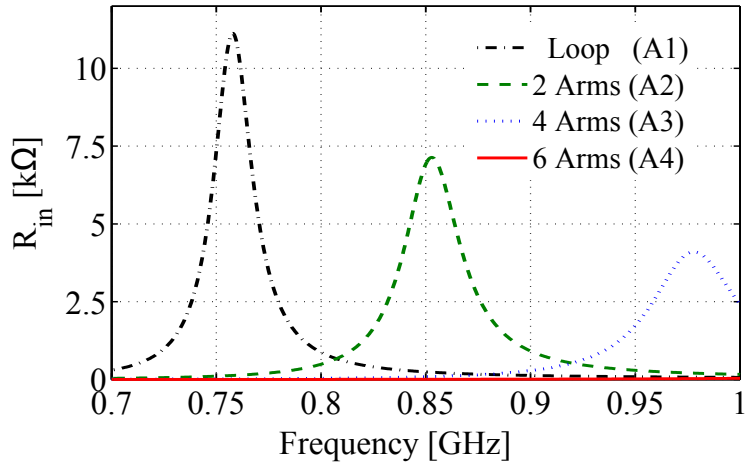
4.4.1 Impedance Tuning Capability

To verify the tuning capability of the suggested antenna configuration, the antenna input impedance and the radiation efficiency are simulated using CST Microwave Studio for the antennas A1, A2, A3 and A4. Figure 4.15 shows the real (a) and imaginary (b) parts of the input impedance as well as the radiation efficiency (c) as a function of frequency for the four simulated antennas. We see, as we expected, that the inner loop dominates the impedance and we see that the resonance shifts toward higher frequencies with the addition of more internal arms. Since the outer arms do draw currents but these currents are much smaller than those in the inner loop, we only observe a slight variation in the radiation efficiency. This demonstrates the impedance-tuning ability of the suggested compact-modified-loop antenna by tuning the variables E_i and W_i, where $i = 1, 2, \cdots, n$, without deteriorating the radiation efficiency of the antenna much. The in-between arms are needed to fine tune the required input impedance.
Table 4.4 summarizes the input impedances and the radiation efficiencies of the four simulated antennas at the design frequency of 868 MHz.

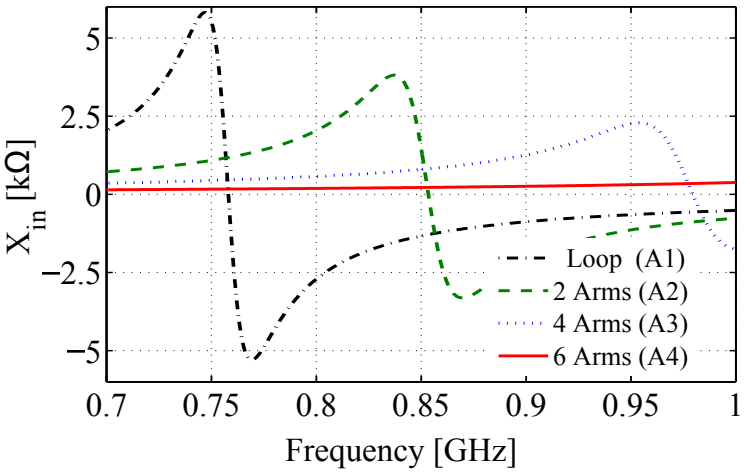
Table 4.4. Simulated input impedance and radiation efficiency at an operating frequency of 868 MHz.

Antenna	A1	A2	A3	A4
$R_{in}(\Omega)$	187.2	3945	129.7	7.08
X_{in} (Ω)	-112.4	-3294	922.3	239.9
$Radiation_{eff}$ (%)	71.51	73.14	74.1	70.7

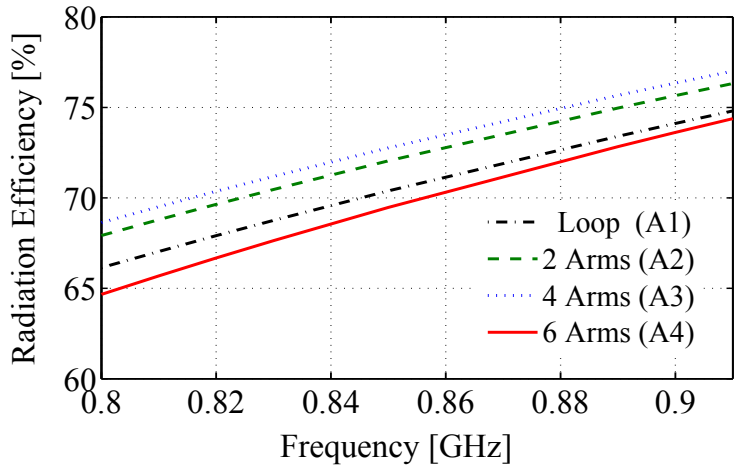
The table shows a large change in the real and imaginary parts of the input impedance when inner arms are added, while the radiation efficiency difference is less than 3.4% over all configurations. In Subsecection 4.4.3 we will further discuss the radiation efficiency simulation results.



(a)



(b)



(c)

Figure 4.15. Real (a) and imaginary (b) parts of the input impedance and the radiation efficiency (c) as a function of frequency of the four simulated antennas.

The main benefits of the suggested novel antenna topology are its compact size ($0.16\lambda \times 0.093\lambda \times 0.0046\lambda$) and the ability to tune its input impedance while maintaining a constant radiation efficiency.

4.4.2 Validation of the Simulation Results

To cross-verify the simulation results two different simulation software packages have been employed: CST Microwave Studio [57] and Empire XCcel [83] that use two different computational methods. CST uses the Finite Integration Technique (FIT) [72] and Empire XCcel uses the Finite Difference Time Domain Method (FDTD) [84]. Antenna A4 (Fig. 4.14(d)) is selected for simulation. The simulated real and imaginary parts of the input impedance are plotted versus frequency in Fig. 4.16. It is shown in the figure that the two different simulation methods converge to almost identical values in the shown frequency range. At a frequency of 868 MHz, the input impedances are found to be: $Z_{in,\,FDTD,\,868\,MHZ} = 8.49 + j237\Omega$ and $Z_{in,\,FIT,\,868\,MHZ} = 7.08 + j239.9\Omega$. The power wave reflection coefficient is calculated using the figure of merit (FoM) introduced in Sec. 1.6. The FoM results in -13.8 dB power reflections at 868 MHz ensuring accurate simulation results.

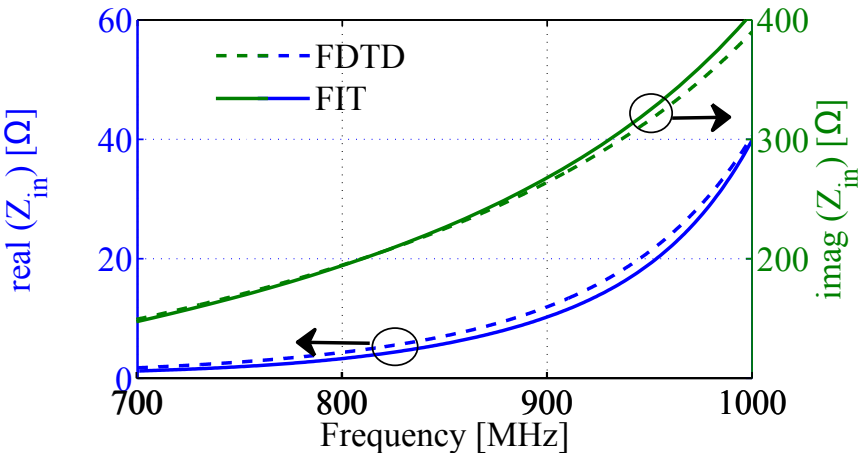


Figure 4.16. Real (a) and imaginary (b) parts of the input impedance as a function of frequency simulated using the Finite Difference Time Domain (FDTD) method and the Finite Integration Technique (FIT).

To verify the simulated results by measurements, a monopole version of the antenna has been fabricated by cutting the antenna through its symmetry plane (see Fig.4.13), and soldering this monopole to a large ground plane as shown in Fig. 4.17. In this way, the antenna can be measured using a single-ended (SMA) connection, which matches with the use of a standard Vector Network Analyser (VNA). The input impedance of the fabricated antenna is measured and compared to the simulated results. Figure 4.18 shows the simulated and measured real (Fig. 4.18(a)) and imaginary (Fig. 4.18(b)) parts

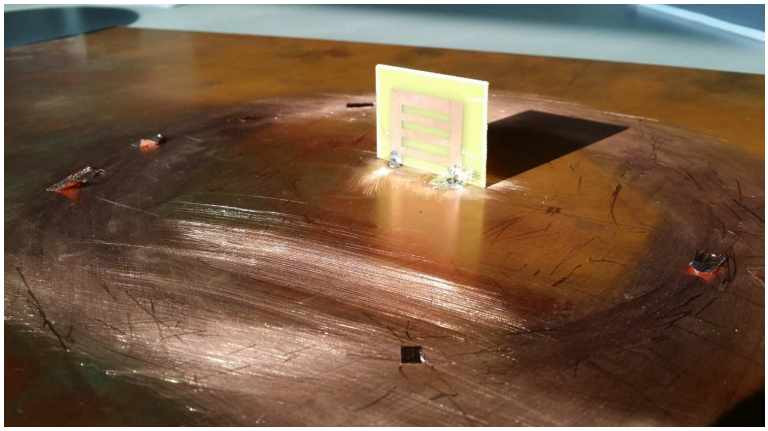


Figure 4.17. Fabricated monopole prototype for impedance measurements.

of the input impedance as a function of frequency. The measured values are multiplied by two to give results that apply for the actual antenna. It is shown in the Figure that the measurement results verify the simulated ones.

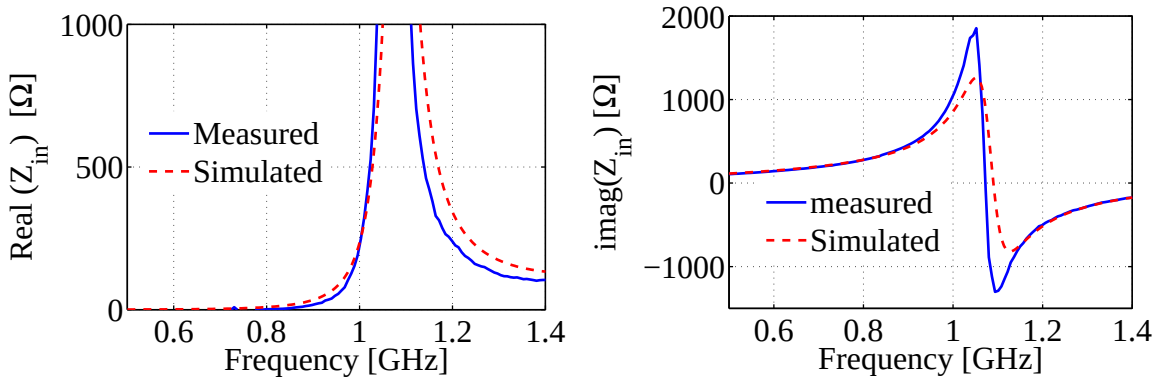


Figure 4.18. Simulated and measured real (a) and imaginary (b) parts of the input impedance of the fabricated prototype antenna versus frequency.

4.4.3 Radiation Efficiency and Radiation Pattern

The antenna radiation efficiency η is defined as [50]:

$$\eta = \frac{R_{rad}}{R_{rad} + R_{loss}}, \tag{4.11}$$

where R_{rad} is the radiation resistance and R_{loss} is the loss resistance.

The Wheeler cap method [85, 86] enables separate measurements of the radiation resistance (R_{rad}) and loss resistance (R_{loss}) by enclosing the antenna with a radiation shield.

The radiation shield is a conducting sphere enclosing the near-field of the antenna at a radius $r = \lambda/2\pi$ [85] as shown in Fig. 4.19(b).
The input impedance of the antenna (Z_{in}) is first measured as indicated in the previous section and illustrated in Fig. 4.19(a).

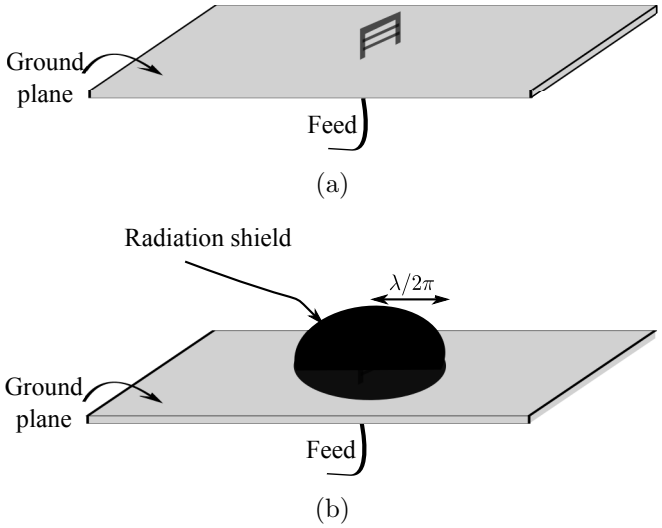


Figure 4.19. Input impedance measurement setup (a) and Wheeler cap measurement setup (b).

The real part of the input impedance is expressed as:

$$Re\{Z_{in}\} = R_{rad} + R_{loss}. \tag{4.12}$$

By placing the radiation shield over the antenna, the radiation resistance (R_{rad}) can be eliminated and consequently the Wheeler cap input impedance (Z_{wc}) can be measured where the real part is expressed as:

$$Re\{Z_{wc}\} = R_{loss}. \tag{4.13}$$

Substituting Eqs. (4.13) and (4.12) in Eq. (4.11), the antenna radiation efficiency is expressed as:

$$\eta = \frac{R_{rad}}{R_{rad} + R_{loss}} = \frac{Re\{Z_{in}\} - Re\{Z_{wc}\}}{Re\{Z_{in}\}} \tag{4.14}$$

The shape of the cap is not critical and need not be spherical [87], [88] and [89]. In this section, a cylindrical cap with a radius of 10.5 cm and a height of 7.5 cm will be used to measure the radiation efficiency of the manufactured prototype. The cylindrical cap

measurement setup is shown in Fig. 4.20. During the measurements, the lid in Fig. 4.20 is placed to enclose the near field of the antenna.

Figure 4.20. Radiation efficiency measurement setup using a cylindrical cap.

The simulated and the measured radiation efficiencies are plotted in Fig. 4.21. It is shown in the figure that the measured results exhibit two major resonances around 1.1 GHz and 1.8 GHz. These resonances are due to the cutoff frequencies of the TM_{01} (1.12 GHz) and TE_{01} (1.78 GHz) propagating modes of the used cylindrical waveguide [80].

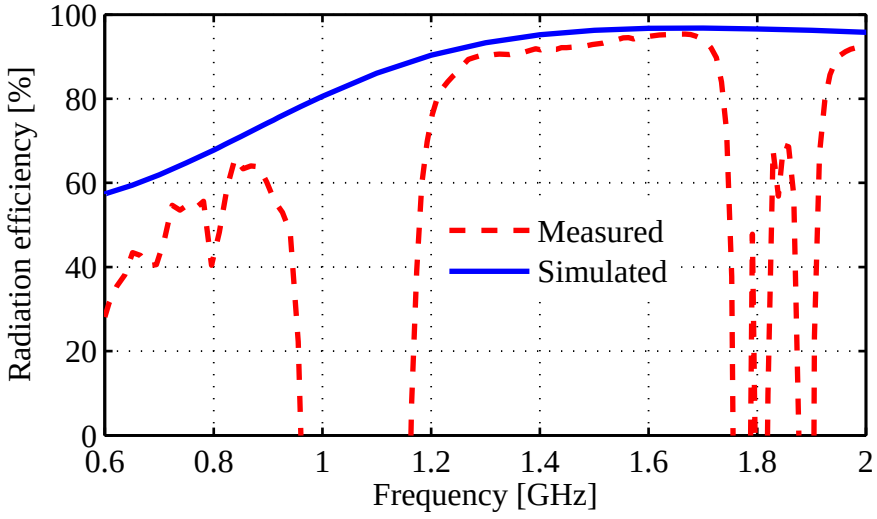


Figure 4.21. Simulated and measured radiation efficiency as a function of frequency using the Finite Integration Technique (simulation) and the Wheeler cap method (measurement).

It should be mentioned that for frequencies below 1 GHz, the height of the cylindrical

cap (7.5 cm) does not satisfy the radian sphere criterion ($r = \lambda/2\pi$) and consequently an extension of the cap height is needed. The cap height is extended by 5 cm using a 5 mm-thick copper plate. The measured radiation efficiency as a function of frequency below 1 GHz using the extended cap is shown in Fig. 4.21. It is clearly indicated in the figure that the deviations between the simulated and the measured radiation efficiencies increase at low frequencies. The deviations are less than 2 % at 1.6 GHz, while they are almost 30% at 0.6 GHz. This behaviour is mainly due to the small value of the input resistance at low frequencies as shown in Fig. 4.16. At low frequencies ($\leq$ 1GHz) accurate small input resistance measurement is quite a challenge due to the imperfections of the calibration standards used which leads to higher deviations in the measured radiation efficiency [90]. Another source of errors at lower frequencies is the non-perfect galvanic contact of the extension of the cylindrical cap.
Excluding the resonant modes of the cylindrical cap, the radiation efficiency has been measured between 1.3 GHz and 1.6 GHz for antennas A1, A2, and A4 (see Fig. 4.14). The simulated and the measured radiation efficiency as a function of frequency are shown in Fig. 4.22. It is indicated in the figure that the deviations between the simulated and the measured radiation efficiency do not exceed 5 % in the measured frequency band. The simulated radiation efficiency has been verified also by employing full-wave Wheeler cap simulations [91].

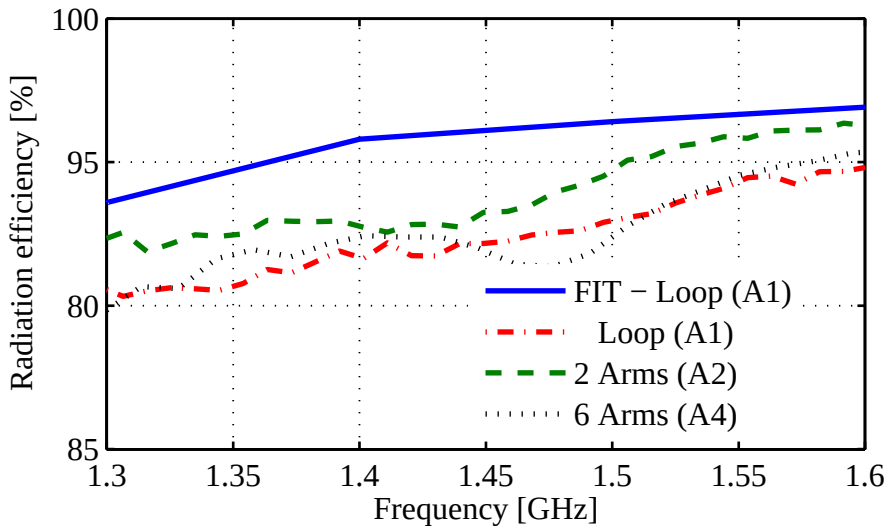


Figure 4.22. Simulated and measured radiation efficiency as a function of frequency for antennas A1, A2 and A4.

The antenna radiation efficiency reaches 70.7 % at the design frequency (868 MHz), see Fig. 4.21. This radiation efficiency can be increased by using a thinner and a lower-loss substrate. The substrate optimization will be presented in the following chapter.

The rectifier is connected on the top side of the antenna, directly to the antenna feeding points, see Fig. 4.23 for the rectifier layout. Figure 4.23 shows the 3D directivity pattern, where the rectifier is placed in the direction of the maximum directivity ($\theta = -90°$, $\phi = 0°$). It is clear from the figure that the layout position of the rectifier does not visibly affect the 3D radiation pattern of the antenna. The directivity D_r and the radiation efficiency $\eta_{radiation}$ determine the antenna gain G_r. $G_r = \eta_{radiation} D_r$, with $D_r = 1.74$ dBi and $\eta_{radiation} = 0.707$, the simulated antenna gain is $G_r = 0.23$ dBi.

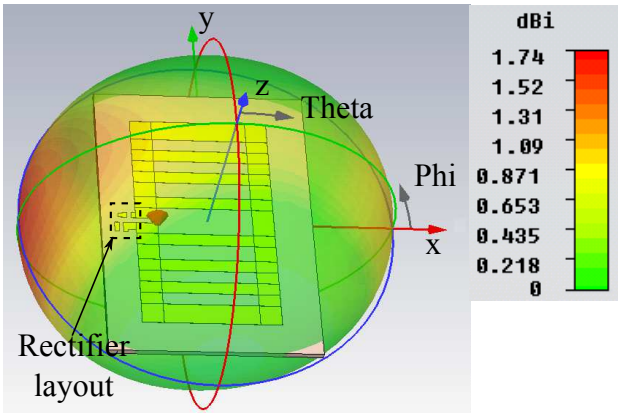


Figure 4.23. 3D directivity pattern at an operating frequency of 868 MHz, simulated using CST Microwave Studio.

As the input impedance of the modified compact loop antenna can be tuned to satisfy the conjugate complex design criterion (i.e. be the complex conjugate of the rectifier input impedance), and the presence of the rectifier does not affect the radiation characteristics of the antenna, this antenna topology will be used in the following chapter to harvest RF power and convert it to usable DC power.

4.5 Conclusion

In this chapter, new design expressions and updated analytical procedures are introduced to calculate the input impedance of strip dipoles and strip folded dipole antennas. The development is driven by the necessity to match the input impedance of the rectifier in a rectenna for maximum power transfer. Simulation and measurement results have shown that the proposed equations satisfy the Figure of Merit criterion and are easily implementable in standard computing tools. Compared to the state-of-the-art analytical solutions, the validity range for the dipole length is extended by 150 % which allows for the modeling of longer strip dipole antennas.
In addition, a novel and compact modified loop antenna configuration has been intro-

duced. The modification consists of adding additional horizontal arms to the loop antenna, which results in a higher degree of freedom to tune the antenna's input impedance. The main advantage of the introduced antenna topology is that the antenna impedance can be tuned without affecting its radiation efficiency much. The simulation results of the input impedance are validated by measurements.

Chapter five

RF Power Transport Using Conjugate-Matched Rectenna System

In this chapter, the modified compact loop antenna introduced in the previous chapter is tuned to eliminate the commonly used matching network between the antenna and the rectifier, since this antenna will be conjugately matched to the rectifier. Two rectenna prototypes are presented in this chapter. The first prototype consists of a rectenna using commercially available, discrete rectifiers. At an incident RF power level of -20 dBm, the Power Conversion Efficiency (PCE) reaches 33.8 % for a resistive load of 10 kΩ. The PCE increases to 54.7 % for a power level of -10 dBm. Compared to the state-of-the-art, the PCE is at least 10 % higher for the same input power levels.
For the second prototype, an integrated 5 stage cross-connected bridge rectifier with a capacitor bank, implemented in standard 90-nm CMOS technology has been used. The PCE reaches 30 % at an input power level of -20 dBm and it increases to 36.6 % at an input power level of -17 dBm over a 0.33 MΩ resistive load. Measurements show a -27 dBm sensitivity for 1 V DC output. At an input power level of -20 dBm, the peak output voltage is measured at an operating frequency of 855 MHz reaching 1.73 V. The suggested prototype is smaller in size than those reported in the literature.

5.1 Introduction

Figure 5.1 shows the configuration of the wireless power transmission system being used in this chapter.

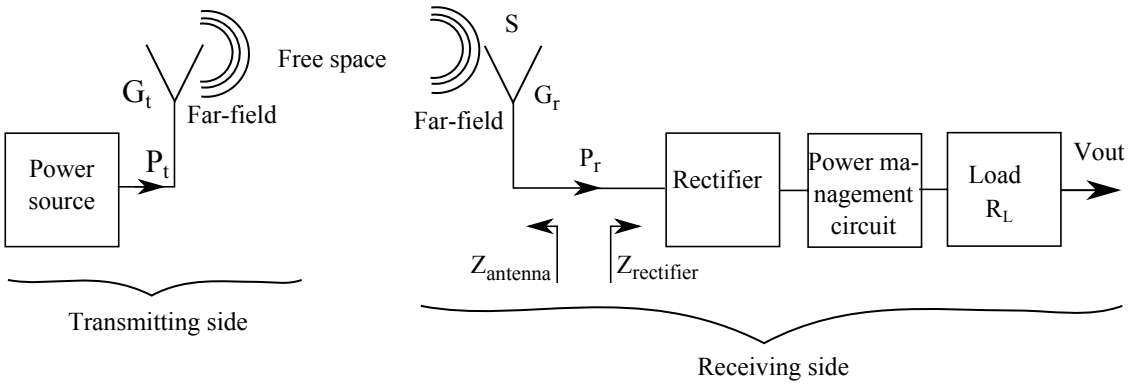


Figure 5.1. Wireless power transmission system.

The system consists of a transmitter that transmits RF power and a receiver that collects part of the transmitted power.
When the transmitter is not dedicated for wireless power transmission, e.g. a TV broadcasting or a GSM base station, the power collection and conversion is defined as wireless power *harvesting* [22], [92]. When the power is intentionally transmitted using a dedicated source, this is defined as wireless power *transport* [93], [94]. For both, the receiving side consists of a *rectenna* (*rec*tifying an*tenna*) which is an antenna connected to a rectifier and a load. The load can be an ultra-low power sensor. A power management circuit between the rectifier and the load is needed to transfer and store the DC power. In order to achieve a compact rectenna system, the commonly employed impedance matching network between the antenna and the rectifier is removed as shown in Fig. 5.1 and the antenna is conjugately matched to the rectifier ($Z_{antenna} = Z^*_{rectifier}$). The compact antenna configuration introduced in the previous chapter is used to capture part of the transmitted RF power.

5.2 Design Methodology

In this section[1], a design methodology is presented to optimize the power transfer between the antenna and the rectifier [95]. The optimal impedance matching network with

[1]This section is based on the work by Hubregt J. Visser that is published in [95]. For the completeness of the work, this section is included in the thesis.

respect to power efficiency, rectifier input voltage maximization and rectifier sensitivity improvement is investigated. It should be mentioned that the impedance matching may be accomplished by a (discrete) lumped-element matching network and by a distributed network integrated into the antenna. Figure 5.2 shows the equivalent circuit model of a generalized rectenna system.

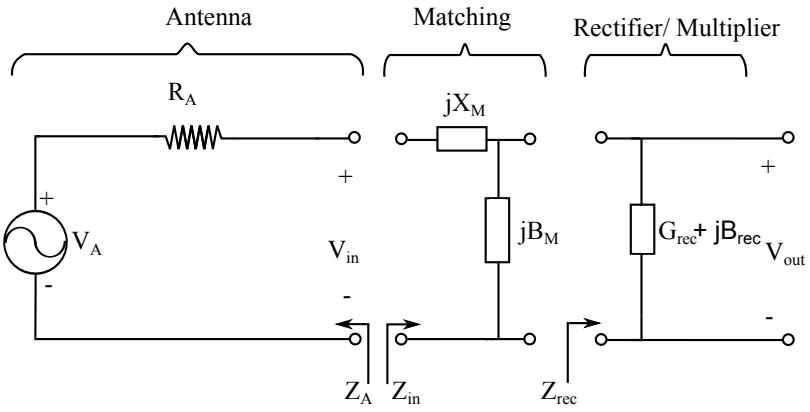


Figure 5.2. Equivalent circuit model of a rectenna system.

The antenna in the figure is represented by its Thévenin equivalent circuit of a voltage source $\mathrm{V_A}$, with an internal resistance R_A. The Thévenin equivalent voltage V_A is:

$$V_A = \sqrt{8 R_A P_{in}}, \tag{5.1}$$

where P_{in} is the maximum power available at the clamps of the antenna.

For the generalized rectenna system shown in Fig. 5.2, an 'L-network' impedance matching is introduced between the antenna and the rectifier/multiplier.
The input power P_{in} at the input of the matching network is described as

$$P_{in} = \frac{1}{2} |V_{in}|^2 \, \Re \{Y_{in}\}, \tag{5.2}$$

where $Y_{in} = \frac{1}{Z_{in}}$ is the input admittance as indicated in Fig. 5.2. The function $\Re\{x\}$ denotes the real part of x. V_{in} is the voltage across the input impedance Z_{in} see Fig. 5.2. V_{in} is calculated as

$$V_{in} = \left(\frac{Z_{in}}{Z_{in} + R_A} \right) V_A = \left(\frac{1}{1 + Y_{in} R_A} \right) V_A. \tag{5.3}$$

Substituting Eq. (5.3) in Eq. (5.2) results in

$$P_{in} = \frac{1}{2}\left|\frac{1}{1+Y_{in}R_A}\right|^2 V_A^2 \Re\left\{Y_{in}\right\}. \tag{5.4}$$

The output power P_{out} across the rectifier terminals is given by

$$P_{out} = \frac{1}{2}\left|V_{out}\right|^2 \Re\left\{Y_{rec}\right\}, \tag{5.5}$$

where Y_{rec} is the input admittance of the rectifier $Y_{rec} = G_{rec} + jB_{rec} = \frac{1}{Z_{rec}}$ and V_{out} is the output voltage across the rectifier terminals. V_{out} and Z_{rec} are indicated in Fig. 5.2.

5.2.1 Antenna Matched to the Rectifier

The 'L-network' shown in Fig. 5.2 is used to conjugately match the antenna impedance to that of the rectifier. If a lossless network is assumed, the output power P_{out} is equal to input power P_{in} and, consequently, the voltage gain G_v is given by

$$G_v = \frac{|V_{out}|}{|V_A|} = \frac{1}{2}\sqrt{\frac{\Re\left\{Y_{in}\right\}}{\Re\left\{Y_{rec}\right\}}}. \tag{5.6}$$

The input impedance Z_{in} is calculated using the network model shown in Fig. 5.2 as

$$\begin{aligned} Z_{in} &= jX_M + \frac{1}{G_{rec} + j\left(B_M + B_{rec}\right)} \\ &= \frac{G_{rec}}{G_{rec}^2 + \left(B_M + B_{rec}\right)^2} + j\left(X_M - \frac{B_M + B_{rec}}{G_{rec}^2 + \left(B_M + B_{rec}\right)^2}\right). \end{aligned} \tag{5.7}$$

In order to conjugately match the antenna to the rectifier, the following conditions apply

$$R_A = \frac{G_{rec}}{G_{rec}^2 + \left(B_M + B_{rec}\right)^2}, \tag{5.8}$$

$$X_M = \frac{B_M + B_{rec}}{G_{rec}^2 + \left(B_M + B_{rec}\right)^2}. \tag{5.9}$$

From Eqs. (5.7) and (5.9), the real part of the input admittance is calculated as

$$\Re\{Y_{in}\} = \frac{G_{rec}^2 + (B_M + B_{rec})^2}{G_{rec}}, \tag{5.10}$$

and, since $Y_{rec} = G_{rec} + jB_{rec}$, the real part of the rectifier's admittance is

$$\Re\{Y_{rec}\} = G_{rec}. \tag{5.11}$$

Substituting Eq. (5.10) and Eq. (5.11) in Eq. (5.6) leads to the voltage gain

$$G_v = \frac{|V_{out}|}{|V_A|} = \frac{1}{2}\sqrt{\frac{\Re\{Y_{in}\}}{\Re\{Y_{rec}\}}} = \frac{1}{2}\sqrt{1 + \left(\frac{B_M + B_{rec}}{G_{rec}}\right)^2} = \frac{1}{2}\sqrt{1 + (Q)^2}, \tag{5.12}$$

where $Q = \frac{B_M + B_{rec}}{G_{rec}}$ is the quality factor at the output of the matching circuit.

G_{rec} and B_{rec} are defined by the rectifier's impedance. Eq. (5.12) leaves the impression that, by a proper choice of the matching network susceptance B_M, one can choose the output voltage V_{out} and then design the antenna having an impedance R_A and take X_M as dictated by Eq. (5.8) and Eq. (5.9), respectively.

The error made in this reasoning is that it is implicitly assumed that V_A and R_A can be chosen independently, while in fact they are related through the available power from the antenna as specified in Eq. (5.1).

5.2.2 Maximizing the Output Voltage

Assuming a lossless network, one can express the available power from the antenna by using Eq. (5.1)

$$P_{in} = \frac{V_A^2}{8R_A}. \tag{5.13}$$

By the substitution of Eq. (5.8) in Eq. (5.13), V_A is expressed as

$$V_A = \sqrt{8R_A P_{in}} = 2\sqrt{2P_{in}\frac{G_{rec}}{G_{rec}^2 + (B_M + B_{rec})^2}}. \tag{5.14}$$

The output voltage is calculated by substituting Eq. (5.14) in Eq. (5.12), which results in

$$V_{out} = G_v V_A = \sqrt{\frac{2P_{in}}{G_{rec}}}. \tag{5.15}$$

As described in Eq. (5.15), the output voltage is thus dictated by the available power from the antenna and the real part of the input admittance of the rectifer/multiplier.

G_{rec} is the real part of the rectifier admittance as shown in Fig. 5.2.

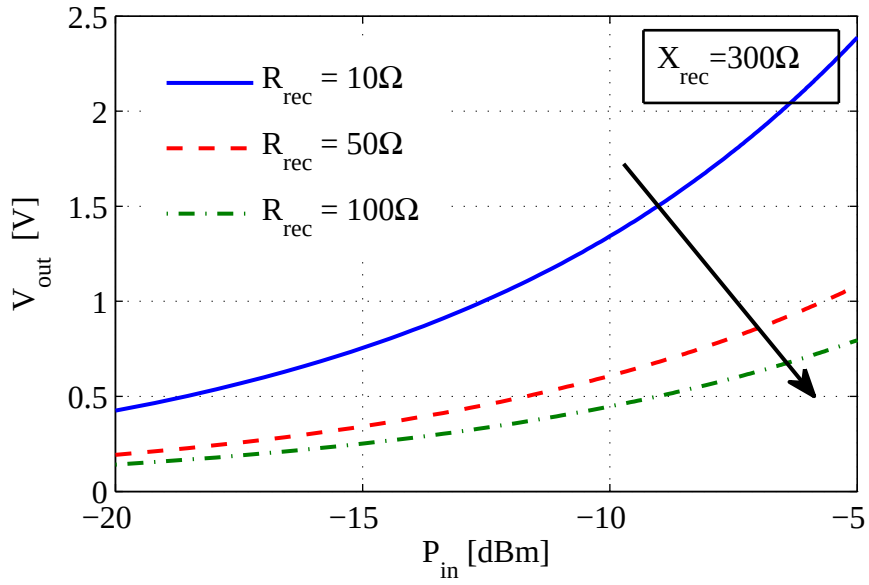
$$\begin{aligned} G_{rec} &= \Re\{Y_{rec}\} \\ &= \frac{R_{rec}}{R_{rec}^2 + X_{rec}^2}. \end{aligned} \tag{5.16}$$

After substituting Eq. (5.16) in Eq. (5.15), the output voltage is described in terms of the input impedance of the rectifier as follows:

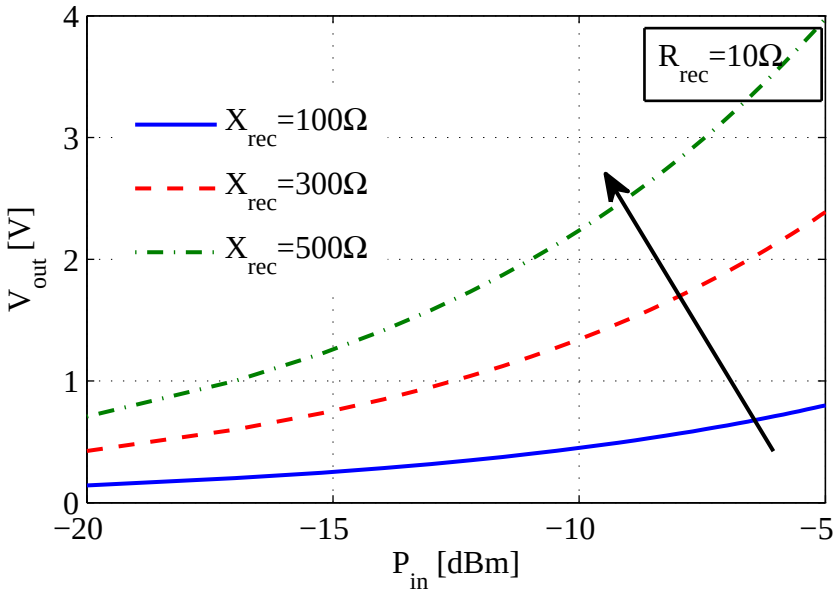
$$V_{out} = \sqrt{2P_{in}\frac{R_{rec}^2 + X_{rec}^2}{R_{rec}}}. \tag{5.17}$$

To maximize the output voltage V_{out} (i.e. the sensitivity) and the power transfer between the antenna and the rectifier, Eq. (5.17) should be satisfied. The real part of the input impedance should be as high as possible while the imaginary part should be as low as possible or the other way around. Since most Schottky diodes show an input impedance having a low resistance in series with a high capacitance we choose for the latter method. Figure 5.3 shows the DC output voltage for different input power levels for different rectifier impedances following Eq. (5.17). Figure 5.3(a) shows the output voltage for different values of the real part of the rectifier impedance while the rectifier reactance is set to 300 Ω. Figure 5.3(b) shows the output voltage for different values of the imaginary part of the rectifier impedance while the rectifier resistance is set to 10 Ω.

It is clear from the figure that when $X_{rec} \gg R_{rec}$, the voltage decreases with increasing real part of the rectifier's input impedance. A voltage of 0.5 V is generated at an available power level of -9 dBm (dotted-dashed curve) when R_{rec} is set to 100 Ω. The same output voltage is generated at an available power level of -18.5 dBm (solid curve) when R_{rec} is decreased to 10 Ω, which results in a 8.5 dB sensitivity improvement. Figure 5.3(b) clearly indicates that the output voltage increases with increasing reactive part of the input impedance when the real part is set to 10 Ω.



(a)



(b)

Figure 5.3. Output voltage for different input power levels P_{in} for different rectifier impedances. (a) $X_{rec} = 300\Omega$, (b) $R_{rec} = 10\Omega$.

A voltage of 0.7 V is generated at an available power level of -6 dBm (solid curve) when X_{rec} is set to 100 Ω. The same output voltage is generated at an available power level of -20 dBm (dotted-dashed curve) when X_{rec} is increased to 500 Ω, which results in a 14 dB sensitivity improvement. This knowledge will be used in Sec. 5.4 where a 5 stages rectifier implemented in TSMC 90-nm CMOS technology with an input impedance of

$Z_{in} = 7.6 + j418$ has been designed, realized and measured.

Through the presented analysis and our results for a compact loop antenna, it has been shown that the steps to maximize the output voltage and thus improve the rectifier's sensitivity for a given low power level are:

- conjugate match the antenna to the rectifier;
- design the rectifier circuit for a minimum real part and maximum absolute value of the imaginary part $|X_{rec}| \gg R_{rec}$;
- use antenna miniaturization up to a level that the radiation resistance still dominates the loss resistance;
- use a small loop antenna as a basis.

5.3 Power Transport Using Commercially Available Diodes

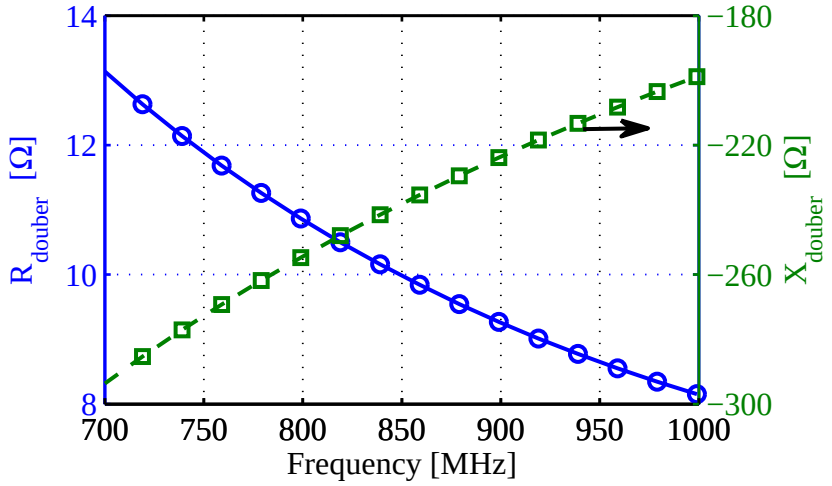
In this section, the modified loop antenna introduced in the previous chapter is combined with commercially available diodes. A voltage doubler rectifier is used. The output voltage is measured for different input power levels at an operating frequency of 868 MHz.

5.3.1 Rectenna Design

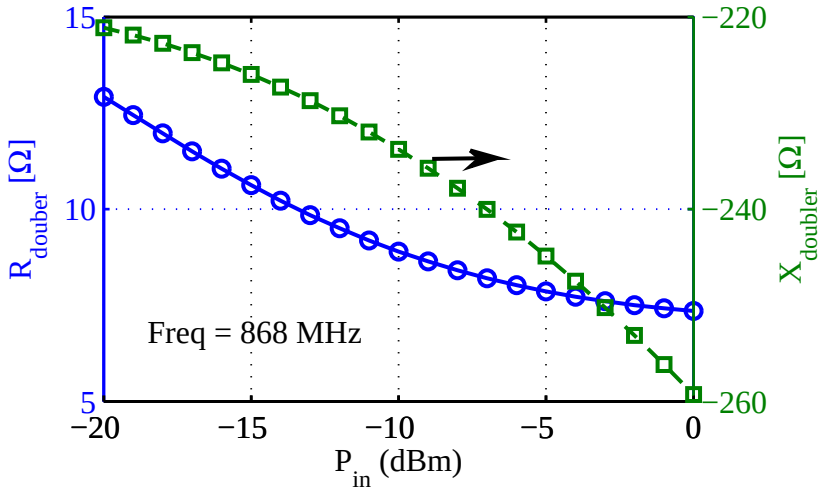
The design frequency and the design input power level have been investigated in Sec. 2.7.1 and Sec. 2.7.2, respectively. The output voltage and the input impedance of a single rectifier (Schottky diode) have been investigated in [41] and have been presented in Chapter 2. In this section we will investigate the input impedance of the voltage doubler (see Fig. 3.21) as a function of frequency for different RF input power levels. The main reason to choose a voltage doubler is that it provides a higher DC output voltage than the single rectifier and it eliminates the need for a return-current coil. In addition, it provides a compact size due to the commercially available double-packaged-diodes. ADS harmonic balance and Momentum [44] have been used for simulations.

For a voltage doubler circuit made with an Avago HSMS-2852 [43] Schottky-diode pair, the input impedance is calculated as a function of frequency and as a function of input power level. Figures 5.4(a) and 5.4(b) show the real and imaginary parts of the input impedance as a function of frequency and as a function of input power level, respectively. For Fig. 5.4(a), the input power level is set to -10 dBm while for Fig. 5.4(b) the operating frequency is 868 MHz. It is observed in the figure that the

impedance of the voltage doubler changes as a function of frequency and as a function of input power level. At the design frequency (868 MHz) and input power level (-10 dBm) the input impedance of the voltage doubler is $Z_{doubler} = 8.9 - j234\,\Omega$.



(a)



(b)

Figure 5.4. Simulated real and imaginary parts of the input impedance of the voltage doubler based on the AVAGO HSMS-2852 as a function of frequency for an input power level of -10 dBm (a) and as a function of input power level at an operating frequency of 868 MHz (b).

The modified compact loop antenna introduced in the previous chapter is tuned, so that its input impedance is equal to the conjugate complex of that of the voltage

doubler, at an input power level of -10 dBm, and an operating frequency of 868 MHz. Figure 5.5 shows the real (a) and imaginary parts of the input impedance as a function of frequency for the voltage doubler and for antenna A4 introduced in Sec. 4.4. It is shown in the figure that the real parts of the input impedance for the antenna and the rectifier intersect around 868 MHz. The imaginary parts of the input impedance for the antenna and the voltage doubler are plotted in Fig. 5.5(b). It is indicated in the figure that $\mathrm{imag}(\mathrm{Z_{rectifer}}) = -\mathrm{imag}(\mathrm{Z_{antenna}})$ around 868 MHz.

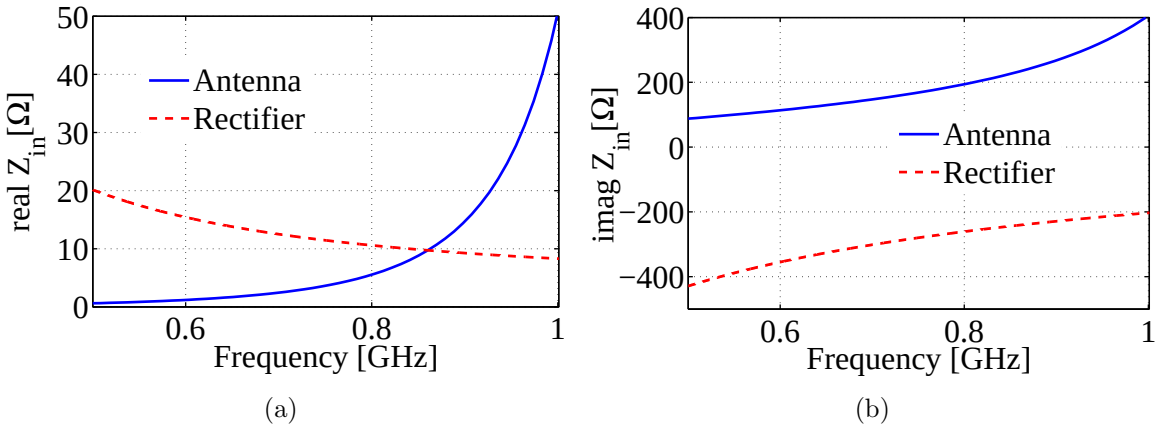


Figure 5.5. Simulated real (a) and imaginary (b) parts of the input impedance of the voltage doubler as well as the antenna as a function of frequency, for P_{in}= -10 dBm.

To investigate the complex conjugate matching, the power wave reflection coefficient concept [34] [51] introduced in Sec. 1.6 is used to calculate the power reflection coefficient between the input impedance of the rectifier and the input impedance of the antenna.

$$\mathrm{FoM} = 20\log\left|\frac{\mathrm{Z_{rectifier}} - \mathrm{Z^{*}_{antenna}}}{\mathrm{Z_{rectifier}} + \mathrm{Z_{antenna}}}\right|, \tag{5.18}$$

where * denotes the complex conjugate of the input impedance. Figure 5.6 shows the power wave reflection coefficient (FoM) as a function of frequency calculated using Eq. 5.18. The figure shows that the antenna is matched to the rectifier with $S_{11} = -16$ dB at the design frequency 868 MHz.
At the design frequency (868 MHz) and at the design input power level (-10 dBm), the input impedance of the voltage doubler is conjugately matched to that of the antenna, which results in a power wave reflection coefficients of -16 dB as indicated in Fig. 5.6. A first rectenna prototype is realized on a 1.6 mm thick FR4 substrate. A photograph of the fabricated rectenna is shown in Fig. 5.7.

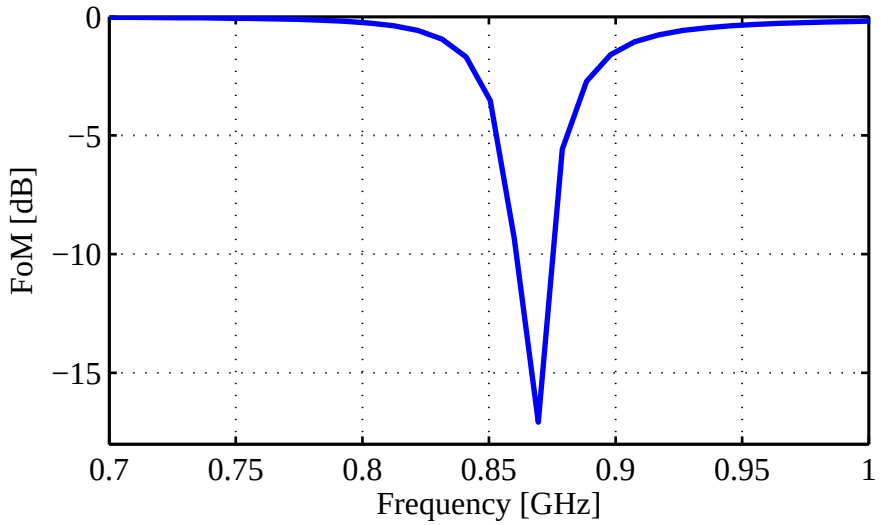


Figure 5.6. Power wave reflection coefficient as a function of frequency calculated using the input impedance of the voltage doubler and the input input impedance of the antenna.

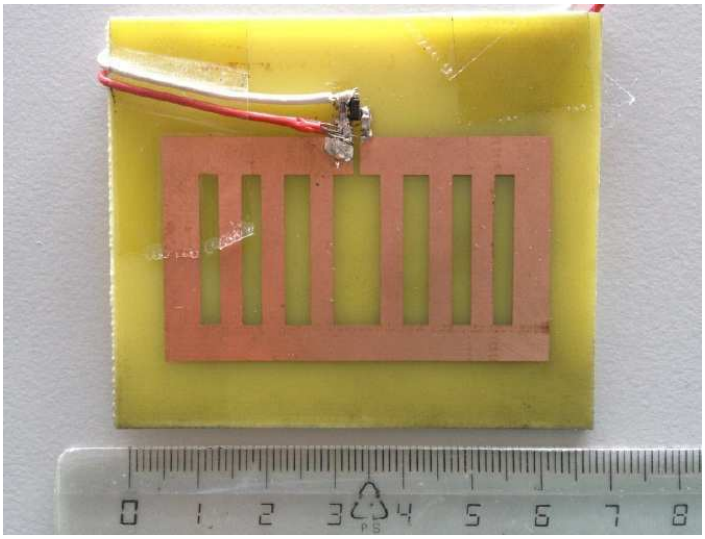


Figure 5.7. Photograph of the fabricated rectenna.

5.3.2 RF Power Transport Measurement Results

For RF power transport, a system as shown in Fig. 3.1 is being used. In this section, we describe how the rectenna is used for wireless RF power transfer. The power transfer measurement setup is shown in Fig. 5.8. The setup is calibrated using two antennas with a known gain, G_r and G_t (see Fig. 5.8), at 868 MHz separated by a distance R.

R is chosen to ensure that far-field conditions are satisfied for both antennas. The power density (S) at the receiving side is dictated by the effective isotropic radiated power, $EIRP = P_t G_t$, where P_t is the transmitted power and G_t is the transmit antenna gain, and by the distance R between the transmitting and the receiving antennas.

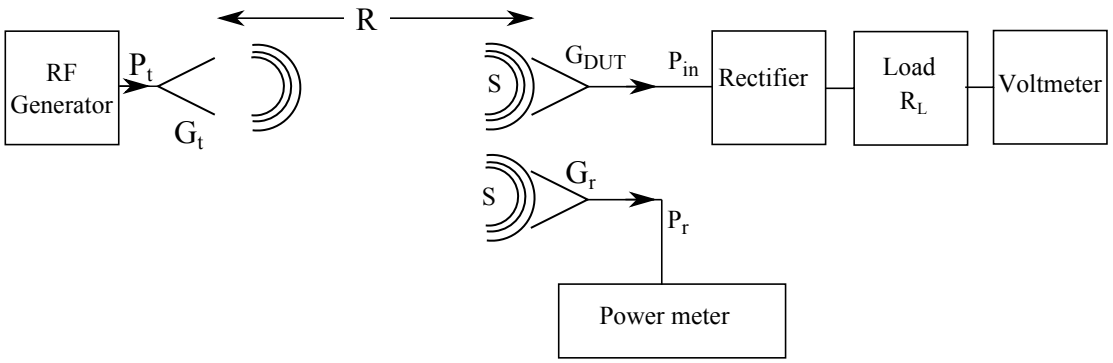


Figure 5.8. Power transfer measurement system.

Using the same procedures as introduced in Sec. 3.7.2 in Chapter 3, the received power at the terminals of the power meter is expressed as

$$P_r = P_t G_t G_r \left(\frac{\lambda_0}{4\pi R}\right)^2 = P_t G_t G_r \left(\frac{c}{4\pi f_0 R}\right)^2, \tag{5.19}$$

where c is the free-space speed of light and f_0 is the operating frequency of the transmitted signal. A power meter is used to measure the received RF power P_r. The measured power P_r is compared to the theoretically calculated value based on Eq. (5.19). The measured power appeared to be within ± 0.5 dB agreement with the theoretically calculated power, which ensures an accurate measurement setup. The rectenna is then placed at the position of the test receiving antenna for wireless power transport. The antenna gain of the rectenna is denoted by G_{DUT} as shown in Fig. 5.8. The input power level Pin at the clamps of the rectifier (see Fig. 5.8) is then calculated as

$$P_{in} = \frac{P_r G_{DUT}}{G_r}. \tag{5.20}$$

The receiving rectenna system is connected to a variable load resistor R_L. The DC output voltage is measured using a voltmeter for different load resistances and for different input power levels using a voltmeter.

The antenna of the receiving rectenna system shown in Fig. 5.8 consists of the modified loop antenna shown in Fig. 5.7. This antenna is connected to the rectifier having a variable load resistor R_L. The DC output voltage is measured for different load resistances

at three different input power levels -20 dBm, -10 dBm and -5 dBm calculated according to Eq. (5.20).

The output voltage is first measured as a function of frequency at an input power level P_{in} (see Fig.5.8) of -10 dBm where a 1 MΩ load resistance is used to terminate the rectenna. Figure 5.9 shows the measured output voltage versus frequency. It is clear from the measurement results that the system output voltage peaks at 868 MHz and reaches 1.36 V. Centered at 868 MHz, the -3dB voltage bandwidth is approximately 30 MHz as indicated in the same figure.

The Power Conversion Efficiency (PCE) is calculated as:

$$\mathrm{PCE}\,(\%) = \frac{P_{load}}{P_{in}} = \frac{V_{out}^2}{R_L}\frac{1}{P_{in}}. \tag{5.21}$$

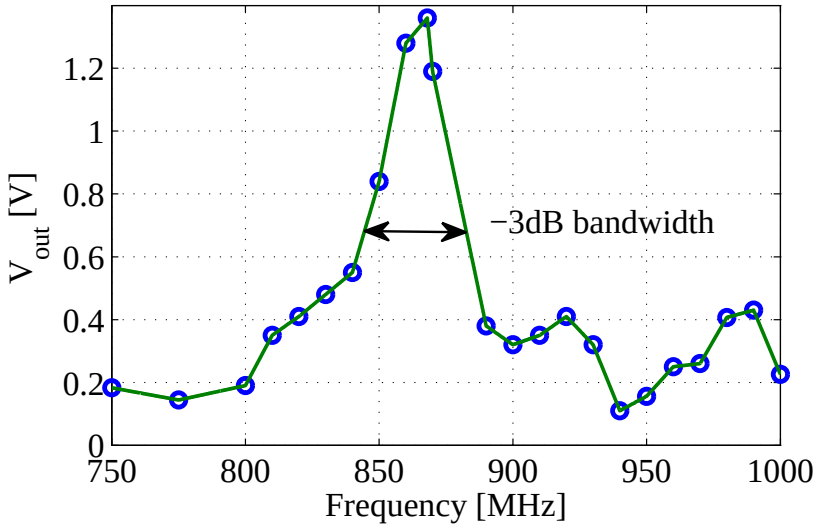


Figure 5.9. Measured output voltage (V_{out}) over a 1 MΩ load resistance as a function of frequency at an input power level P_{in} of -10 dBm.

To measure the power conversion efficiency (PCE), the output voltage is measured over different load resistances and for different input power levels; the results are shown in Fig. 5.10. Over a 10 kΩ resistive load, the DC output voltage reaches 0.19 V at an input power level of -20 dBm. V_{out} increases to 0.76 V and 1.41 V at an input power level of -10 dBm and -5 dBm, respectively.

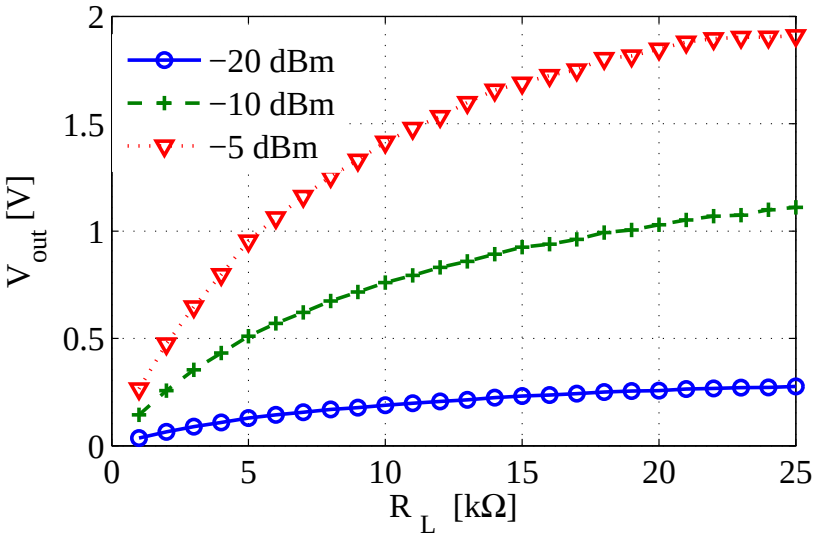


Figure 5.10. Measured output voltage (V_{out}) for different input power levels (P_{in}) versus R_L.

Figure 5.11 shows the measured PCE as a function of R_L for different input power levels (P_{in}) at the design frequency, 868 MHz.

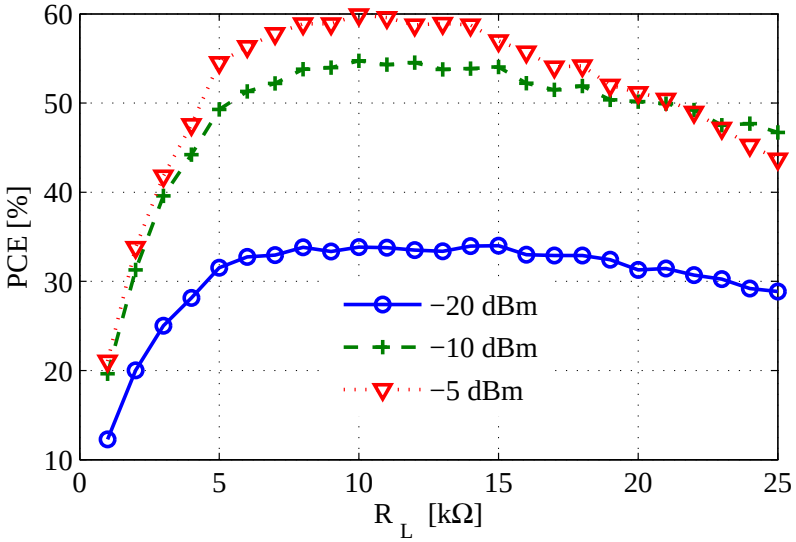


Figure 5.11. Measured power conversion efficiency (PCE) for different input power levels (P_{in}) versus R_L.

At an input power level of -20 dBm, the PCE reaches 33.8 % for a resistive load for 10 kΩ. For the same load resistance, the PCE increases to 54.7 % and 60 % at an input power level of -10 dBm, and -5 dBm, respectively.

Table 5.1 gives a short overview of the state of the art reported for rectennas in the literature that use commercially available rectifiers for RF to DC power conversion.

Table 5.1. Overview of rectenna performances from literature.

Reference	Pin [dBm]	Frequency [MHz]	Load R_L [Ω]	$\frac{Size}{\lambda^2}$	Power Conversion Efficiency (PCE)
[19]	-10 -20	1800	2.4×10^3	-	38*% 15*%
[67]	-10	830	1×10^7	0.027	44 %
[96]	-20	550	2×10^3	0.035	18* %
[20]	0*	1960	460	0.192	54 %
This work	-10 -20	868	10×10^3	0.028	54.7% 33.8%

*calculated from figure

Compared to [67], where a 4 stage rectifier has been used, the presented rectenna in this paper is more efficient at the same input power level and has approximately the same size. At an input power level of -20 dBm, the measured PCE in [96] is 18 %. At the same input power level, the measured PCE of the rectenna presented in this paper reaches 34%. When compared to the measured PCE in [20], our rectenna shows approximately the same performance at an input power level that is 10 times lower. In addition, the rectenna in this chapter is more than five times as compact and operates at a lower frequency. Thus, through a co-design of the antenna and the rectifier, we are able to significantly improve the performance compared to the state of the art.

5.4 Five Stage CMOS Rectifier with a Control Loop

In the previous section a voltage doubler using commercially available diodes has been used to convert the RF incident power into DC power. To further increase the output voltage, a 5 stage rectifier[2] implemented in TSMC 90-nm CMOS technology is designed and realized and the rectifier output voltage and sensitivity are measured.

5.4.1 Design and Implementation

The CMOS rectifier to be presented has been first introduced in [98] and is optimized to harvest power at low input power levels, around -20 dBm. Fig. 5.12 shows the principle

[2]This section is the result of the cooperation between Holst Centre/imec and Delft University of Technical (TUD). The 5 stage CMOS rectifier with the control loop was designed, developed and realized by Mark Stoopman from TUD. The antenna design as well as the connector layout are realized by Shady Keyrouz from TU/e. The reported results in this section are published in [97].

of the multi-stage RF harvester as introduced in [98].

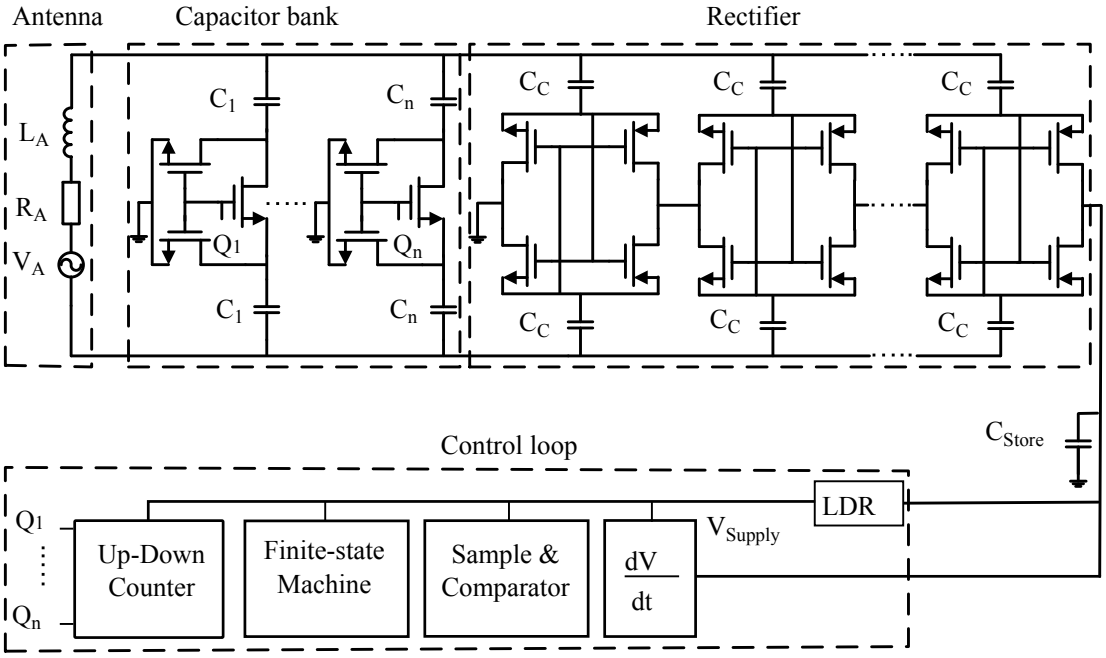


Figure 5.12. Muli-stage RF harvester implemented using CMOS technology.

The harvester consists of a 5 stage cross-connected bridge rectifier. For maximizing the output voltage, the rectifier is connected to a purely capacitive load C_{Store}. A control loop is added to maximize the output voltage on the capacitive load by tuning the reactive part of the rectifier's input impedance such that it is conjugately matched to the antenna. The antenna is connected to a capacitor bank that is controlled by an up-down counter. The 7-bit capacitor bank consists of 128 capacitor switches where C_{max} = 256 fF and C_{min} = 116 fF. By using the control loop, the input impedance of the rectifier as seen from the antenna can be tuned between C_{min} and C_{max} to be conjugately matched to that of the antenna, provided that the real parts of the antenna and rectifier are identical.

The control loop implementation in combination with the multi-stage rectifier is also shown in Fig. 5.12. The harvester initially charges the storage capacitor to the turn-on voltage of the control loop. Then the energy transfer to the off-chip capacitor (C_{Store}) is optimized by maximizing the slope of the load voltage. The slope information is obtained by using a differentiating network. Then the slope information is compared with the previous sample using the sample and comparator stage that determines whether the slope has increased or decreased. Consequently, the finite-state machine determines if the up-down counter should keep counting or change the count direction.

In this realized first prototype, the control loop is implemented off chip by using a micro-

controller. Once the loop is calibrated, it can be turned off so that it is not loading the rectifier for very low input power levels. The input impedance seen from the antenna is found to be $Z_{in,0} = 7.6 - j418\,\Omega$ for C_{min} and $Z_{in,1} = 4.3 - j310\,\Omega$ for C_{max}. This self-calibrating RF rectifier has been implemented in [39] where the rectifier was printed on a FR4 substrate connected to the modified square loop antenna introduced in the previous chapter. In this section, the substrate material and the substrate thickness are optimized for maximum antenna radiation efficiency. A prototype realized on a low-loss and thin substrate will be presented.

5.4.2 Antenna Optimization and Chip Integration

The same antenna topology as introduced in Sec. 4.4 is used for realizing a rectenna based on the just discussed rectifier. In this design the antenna reactance part $X_{Antenna}$, should be between 310 Ω and 418 Ω. Then, the control loop will optimize the rectifier impedance to be conjugately matched to the antenna. The real part should be a fixed value between 4.3 Ω and 7.6 Ω. So an exact, critical value of the reactive part of the antenna impedance is not required. This gives another degree of freedom in maximizing the antenna radiation efficiency.
The antenna radiation efficiency is improved by choosing a thin and a low-loss substrate instead of the so far used lossy FR4 PCB material. Figure 5.13 shows the simulated antenna radiation efficiency versus frequency for different substrate materials for the structure shown in Fig. 5.15.

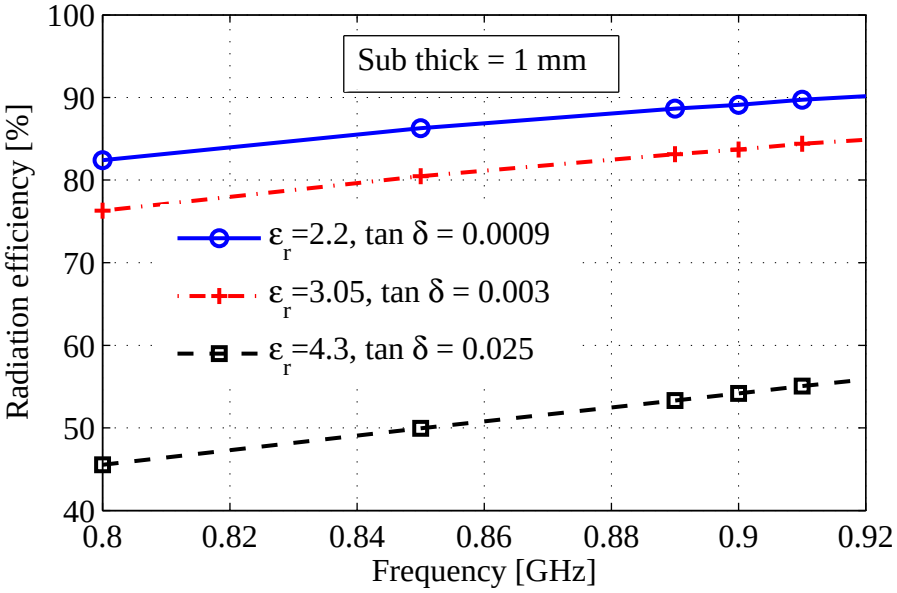


Figure 5.13. Antenna radiation efficiency versus frequency for different substrate materials. Substrate thickness is 1 mm.

In this Figure the substrate height is set to 1mm and the radiation efficiency is simulated for three different substrate materials, FR4 dashed curve ($\epsilon_r = 4.3$, $\tan\delta = 0.025$), GML 1000 dashed-dotted curve ($\epsilon_r = 3.05$, $\tan\delta = 0.003$) and Rogers RT-5880 solid curve ($\epsilon_r = 2.2$, $\tan\delta = 0.0009$). It should be mentioned that the antenna size is kept constant for the three simulations. It is shown in the figure that the radiation efficiency increases when the substrate loss ($\tan\delta$) decreases.

Figure 5.14 shows the antenna radiation efficiency as a function of frequency for varying substrate thickness. In this figure the substrate material is chosen to be GML 1000. The figure shows that a thinner substrate has a limited effect on the radiation efficiency and will result in a slightly higher radiation efficiency. The antenna radiation efficiency is improved by less than 4% when choosing a 0.5 mm thick substrate instead of a 1.5 mm thick substrate.

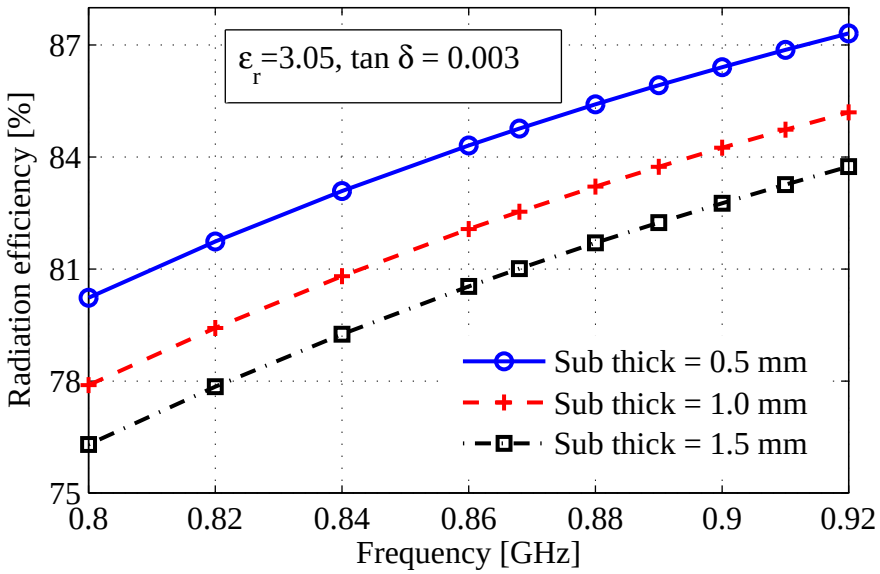


Figure 5.14. Antenna radiation efficiency as a function of frequency for different substrate thickness.

For the prototype realized in this section using the 5 stage CMOS rectifier, a 0.5 mm thick GML 1000 substrate is chosen. At the design frequency, the simulated radiation efficiency is presented in the solid curve of Fig. 5.14. At an operating frequency of 868 MHz, the antenna radiation efficiency is 84.7 %.

The real part of the antenna should be a fixed value between 4.3 Ω and 7.6 Ω. Based on the rectenna design criterion presented in Sec. 5.2.2, the real part of the input impedance of the rectifier is chosen to be 4.4 Ω. After optimizing all the antenna parameters (see Fig. 4.13), an antenna with one short-circuited arm ($n = 1$) having an input impedance

of $Z_{Antenna} = 4.4+j328\Omega$ is chosen. The geometrical parameters of the designed antenna (see Fig. 4.13) are listed in Table 5.2.

Table 5.2. Optimized antenna parameters for the suggested antenna prototype. All dimensions are in mm.

n	W, W_1	g_1, g_2	E_1	L_1	L_2	S_1	S_2	S_3
1	5	3	2	34	35.25	13.75	2	4.5

The front and the back sides of the designed antenna are shown in Fig. 5.15.

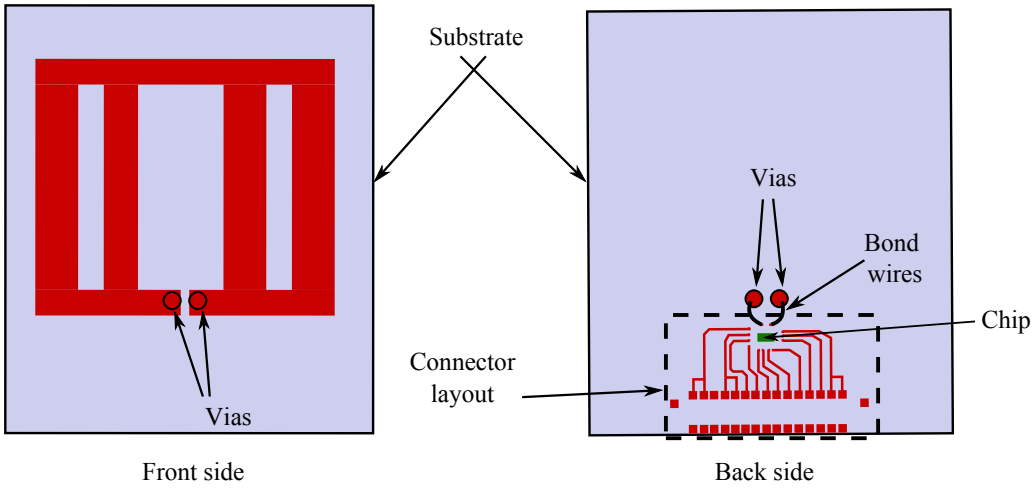


Figure 5.15. Front and back sides of the designed compact antenna structure, including the five stage CMOS chip and the connector layout.

For a proof of principle, the control loop is connected to the chip and antenna using the connector layout shown in Fig. 5.15. The chip is integrated on the backside of the antenna. It should be mentioned that the antenna parameters are optimized in the presence of the connector layout, vias and bond wires. The rectifier inputs are bond-wired to the vias (2 mm in diameter) that connect to the antenna feeding points. The bond wires are modeled using CST Microwave Studio and are having a length of 2.1 mm and a radius of 25 μm.

Due to the self calibrating capability of the circuit and the low real part of the input impedance (R_{in}), the antenna size of the prototype presented in this section is reduced to 21.9 cm^2 compared to 34.3 cm^2 for the first prototype presented in Sec. 5.3.1. Once the external control loop is completely integrated on chip, the connector is no longer required and the antenna board size can be scaled down further without introducing a significant performance loss.

Figure 5.16 shows the simulated 3D directivity pattern of the prototype antenna. The directivity D_r and the radiation efficiency $\eta_{radiation}$ determine the antenna gain G_r. $G_r = \eta_{radiation} D_r$, with $D_r = 1.3$ dBi and $\eta_{radiation} = 0.85$, the simulated antenna gain is $G_r = 0.6$ dBi.

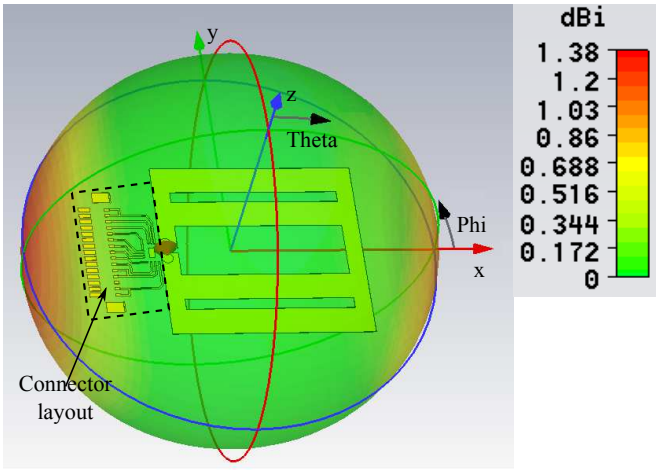


Figure 5.16. 3D directivity pattern of the prototype antenna at an operating frequency of 868 MHz, simulated using CST Microwave Studio.

Figure 5.17 shows a photograph of the manufactured prototype.

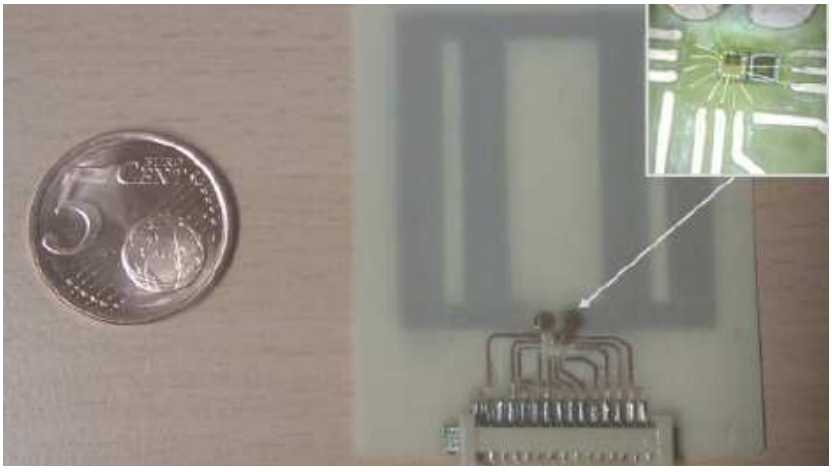


Figure 5.17. Photograph of the manufactured prototype antenna with the integrated circuit. The inset shows the bonded chip in more detail.

5.4.3 Measurement Results

The output voltage and the power conversion efficiency have been measured for different load resistances and for different RF input power levels (P_{in}) at the design frequency (868 MHz).
Figure 5.18 shows the measured DC output voltage (V_{out}) as a function of frequency for a load resistance of 1 MΩ at an input power level of -20 dBm.

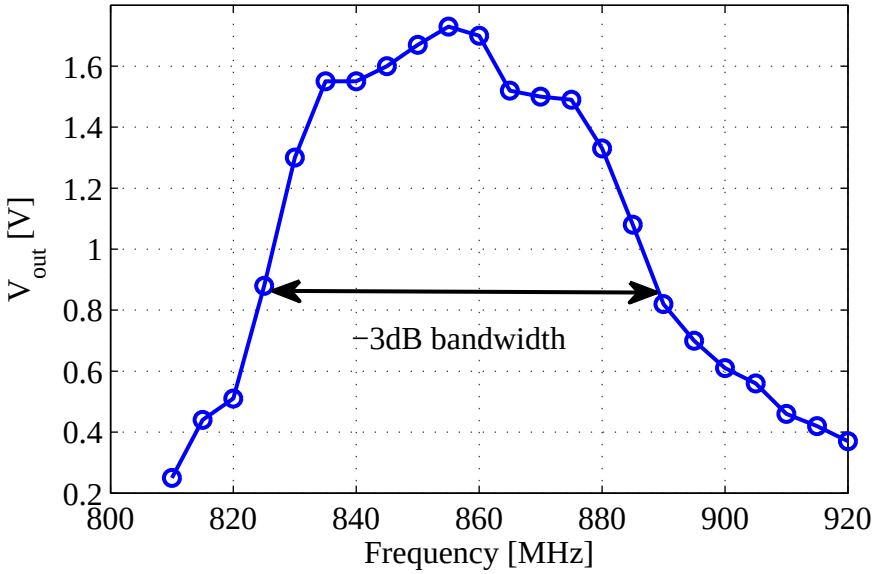


Figure 5.18. Measured output voltage versus frequency for $R_L = 1M\Omega$ and P_{in} = -20 dBm

The peak output voltage is measured at an operating frequency of 855 MHz and reaches 1.73 V. The -3 dB voltage bandwidth is approximately 60 MHz. Compared to the output voltage (measured at Pin = -10 dBm) of the first fabricated prototype, where commercial diodes have been used (see Fig. 5.9), it is clear that this prototype rectenna generates a higher output voltage even at an input power level that is ten times lower (- 20 dBm). In addition, the -3dB voltage bandwidth of this prototype rectenna has almost doubled. This increase in the output voltage was expected and is due to the lower real part and higher imaginary part of the input impedance of the designed CMOS rectifier ($Z_{CMOS-Rectifier} = 4.4 + j328\Omega$) as compared to the input impedance of the discrete diode voltage doubler used for the prototype introduced in the previous section ($Z_{Doubler} = 8.5 + j237\Omega$).

Next, the output voltage has been measured as a function of input power level for different load resistances at 868 MHz; the result is shown in Fig. 5.19. The open circuit output voltage is represented by the dashed-dotted curve in the figure, where the rectenna is able to generate 1 V at an input power level on -27.5 dBm.

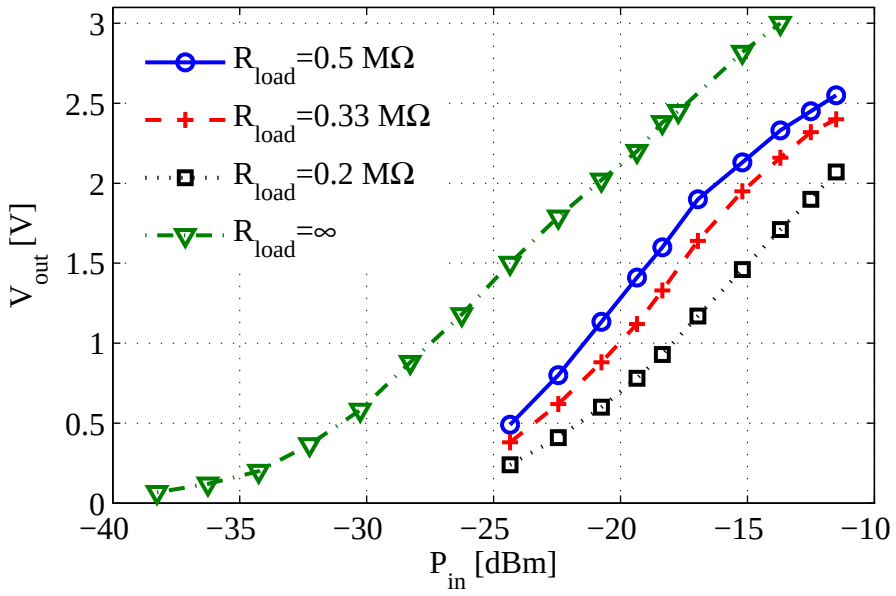


Figure 5.19. Measured output voltage as a function of input power level for different resistive load values.

Based on these voltage levels, and on the simulated antenna gain of 0.6 dBi (D_r = 1.3 dBi, $\eta_{radiation}$ = 0.85), the Power Conversion Efficiency (PCE) is calculated from Eq. (5.21) and plotted in Fig. 5.20 as a function of P_{in} for different load resistance values.

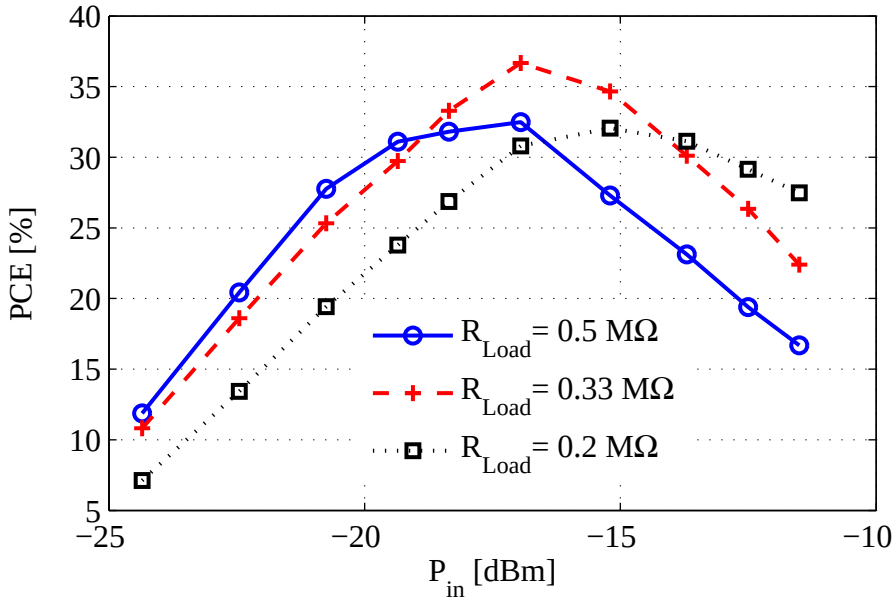


Figure 5.20. Measured power conversion efficiency versus input power level for different resistive load values. Operating frequency = 868 MHz

The power conversion efficiency reaches 30 % at an input power level of -20 dBm and increases to 36.6 % at an input power level of -17 dBm over a 0.33 MΩ resistive load. Compared to the recently reported results in [99, 100], the presented rectenna in this paper has a lower sensitivity and is more compact. Compared to [101] and [102], the proposed rectenna is 70% and 37% more compact respectively. Table 5.3 compares the performances achieved in this work to the state of the art as published in the open literature.

Table 5.3. Overview of rectennas performance with CMOS rectifier in literature.

Ref.	This work	[101]	[102]	[103]
Technology	90 nm	0.18 μm	250 nm	90 nm
Frequency	868 MHz	970 MHz	906 MHz	915 MHz
Sensitivity	-27 dBm @ 1 V	-17.7 dBm @ 0.8 V	-22.6 dBm @ 2 V	-24 dBm @ 1 V
Peak PCE (%)	36.6 @ -17 dBm	37*@ -18.7 dBm	30*@ -8 dBm	11 @ -18.83 dBm
$\frac{Size}{\lambda^2}$	0.018	0.039	0.027	n.a
Measured distance	27 meters @ 1.78 w	n.a.	15meters @ 4 w	n.a.

*calculated from the graph

The suggested co-designed antenna-CMOS rectifier is smaller in size than those reported in other publications. Once the control loop is integrated on chip, the size can be even further reduced. Thus, through the co-design of the 5 stage CMOS rectifier and the antenna and following the design criterion from Sec. 5.2.2 we were able to significantly increase the performance of the rectenna with respect to the DC output voltage, power conversion efficiency, sensitivity and size compared to the state of the art.

5.5 Conclusion

A compact rectenna has been presented in this chapter. The design strategy is based on designing an antenna with a input impedance equal to the complex conjugate of that of the rectifier. This allows removing the matching network, which results in a more compact and efficient rectenna system. Two prototypes have been designed, manufactured, and measured. The first prototype consists of a rectenna that uses commercially available diodes, while in the second prototype a CMOS rectifier is used to convert the RF power into DC power.
In the first prototype, a voltage doubler is used that is conjugately matched to the

antenna. At an available power level of -20 dBm, the PCE reaches 33.8 % for a resistive load of 10 kΩ. The PCE increases to 54.7 % when the RF input power level is increased to -10 dBm. The measured results showed an improved performance compared to recently reported results in literature.

For the second prototype, a 5 stage cross-connected bridge rectifier with a capacitor bank implemented in standard 90-nm CMOS technology has been used. A control loop is used to optimize the matching between the antenna and the rectifier. Compared to the output voltage of the voltage doubler of the first fabricated prototype, where commercially available diodes have been used, the 5 stages bridge rectifier generates a higher output voltage even at an input power level that is ten times lower. In addition, the -3 dB voltage bandwidth of the second prototype is almost doubled. Measurements show a sensitivity of -27 dBm for a 1 V DC output.
At an input power level of -20 dBm, the peak output voltage is measured at an operating frequency of 855 MHz and reaches 1.73 V. The radiation efficiency reaches 30 % at an input power level of -20 dBm and it increases to 36.6 % at an input power level of -17 dBm over a 0.33 MΩ resistive load. Compared to the state of the art published results, a better performance of the rectenna has been achieved with respect to the DC output voltage, power conversion efficiency and sensitivity. Moreover the suggested rectenna is smaller in size than those reported in previous publications.

Chapter six

RF Power Transfer - Complete Rectenna System

In this chapter a complete rectenna system, i.e. the receiving part of the RF power transfer system, is described. The system consists of a signal generator connected to a matching network and a rectifier. The signal generator replaces the receiving antenna to ensure an accurate and repeatable signal injection. A commercially available power management circuit is then used to boost the DC voltage and to store the captured energy in a capacitor. The power management circuit is connected to a dynamic load to emulate different radio systems behaviour. The system's DC average output power and the system's power conversion efficiency are measured and compared to those of a commercially available rectenna system.

6.1 System Concept and Measurement Setup

In this section, the matching network and the voltage doubler introduced in Sec. 3.7.1 are combined with the power management circuit TI BQ25570 [104] and a dynamic load that emulates the behaviour of a radio system. The measurement system[1] is shown in Fig. 6.1.

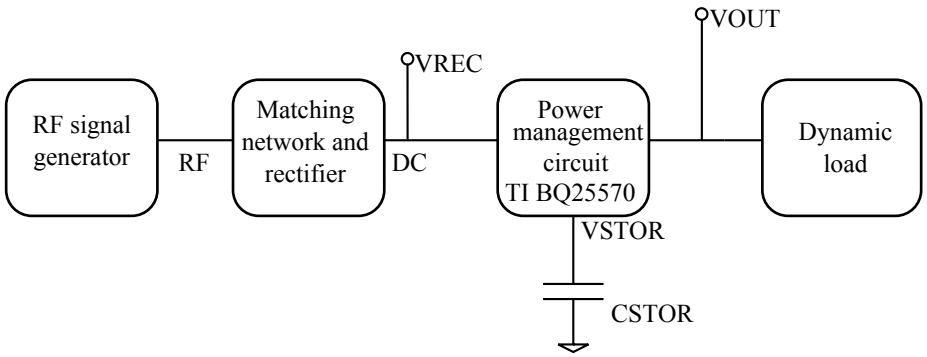


Figure 6.1. Measurement setup using a power management circuit and a dynamic load.

The RF signal generator in this setup replaces the (50Ω) antenna, so that a stable and repeatable measurement setup is created. Since the DC output voltage of the rectifier stage VREC (see Fig. 6.1) is limited and changes with input power level, a power management circuit is needed to store and regulate the system output energy. For our system, the power management circuit TI BQ25570 is used. The connections to the TI BQ25570 are shown in Fig. 6.2. A complete schematic and explanation of the evaluation board used are presented in [105].
The integrated boost controller in the chip amplifies the DC input voltage (VIN_DC=VREC) and controls its output voltage (VSTOR). The boost controller stops when VSTOR reaches a threshold level that is set with the external resistor network shown in Fig. 6.2. The energy is stored in a (super)-capacitor or a rechargeable battery. The integrated buck converter (see Fig. 6.2) is powered from VSTOR and regulates the output voltage to drive different applications. To emulate different radio systems behaviour a dynamic load is connected to the VOUT pin of the evaluation board. This dynamic load will be discussed in more detail in the remainder of this chapter.
The amount of energy stored in a capacitor C, when the voltage increases from V_1 to

[1]The measurement setup as well as the measurement results presented in this chapter were performed at the Holst Centre/imec by Hans Pflug.

V_2 is:

$$W = \frac{1}{2} C \left(V_2^2 - V_1^2 \right). \tag{6.1}$$

The minimum capacitance value, C_{min}, depends on the power drawn by the dynamic load P_L during a discharge time T_p when the capacitor voltage drops from V_2 to V_1 and is given by:

$$C_{min} \geq \frac{P_L T_p}{0.5 \left(V_2^2 - V_1^2 \right)}. \tag{6.2}$$

For example, suppose 60 mW of power is drawn for 40 ms, resulting in a voltage drop from 4.2 V to 1.8 V. This system will have a $C_{min} \geq 333\ \mu F$. To ensure sufficient storage capacitance, a 1000 μF (CSTOR) capacitor is connected to the VSTOR-pin of the TI chip.

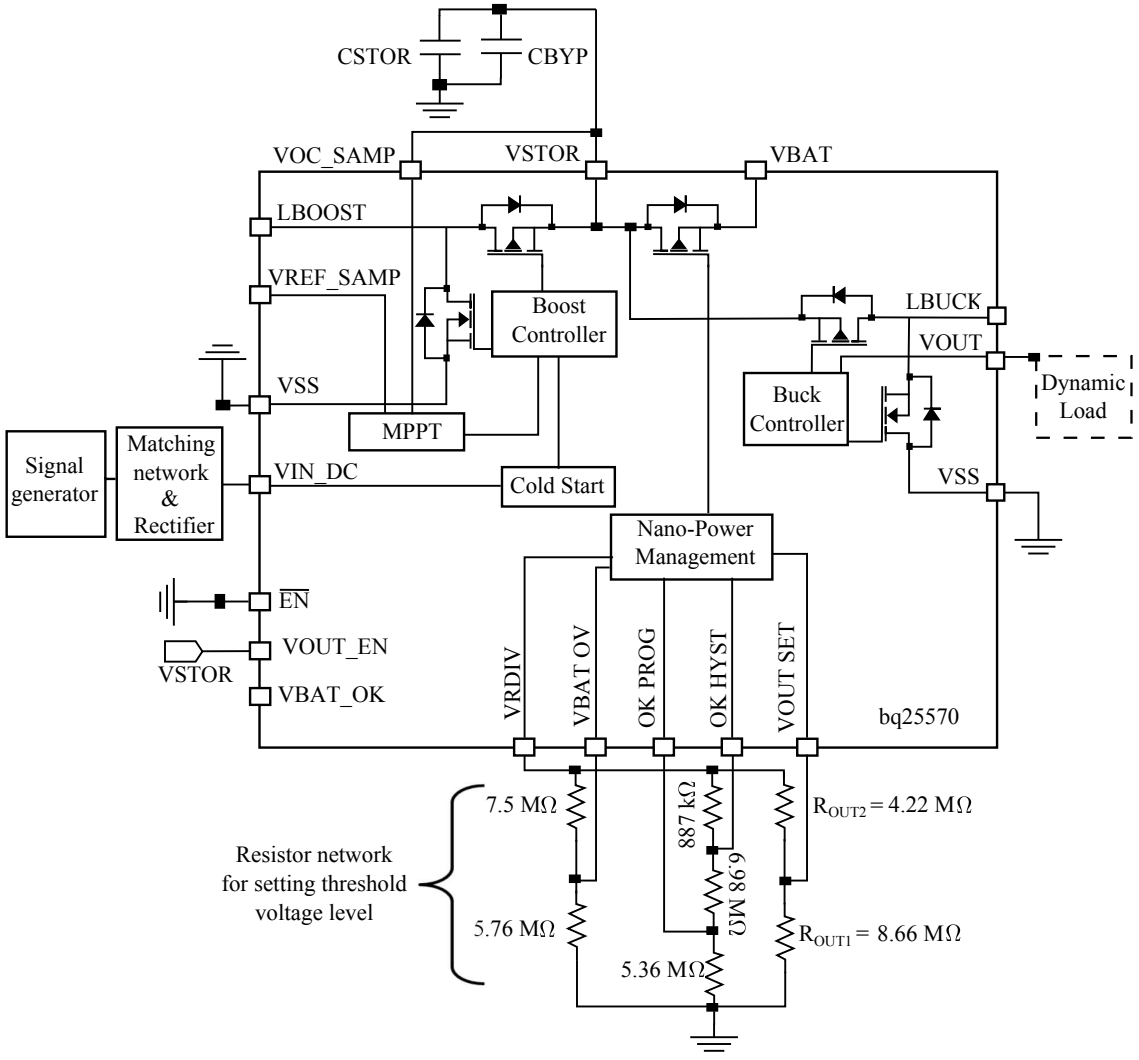


Figure 6.2. Schematic connection to the TI BQ25570 evaluation board.

The integrated boost controller provides 4.2 V at the VSTOR-pin. The buck controller regulates the output voltage VOUT (see Fig. 6.2) at 1.8 V for the chosen values of the resistor network shown in Fig. 6.2. A dynamic load is connected at the VOUT-pin of the TI chip.

The schematic of the dynamic load is shown in Fig. 6.3. It consists of an array of load resistors connected to BJTs that are operated as switches. The switches are controlled by an Arduino Mega board [106] with the ATmega2560 Atmel micro-controller. So, the load resistance and the on-time (T_P) and off-time (T_i) are at the users disposal.

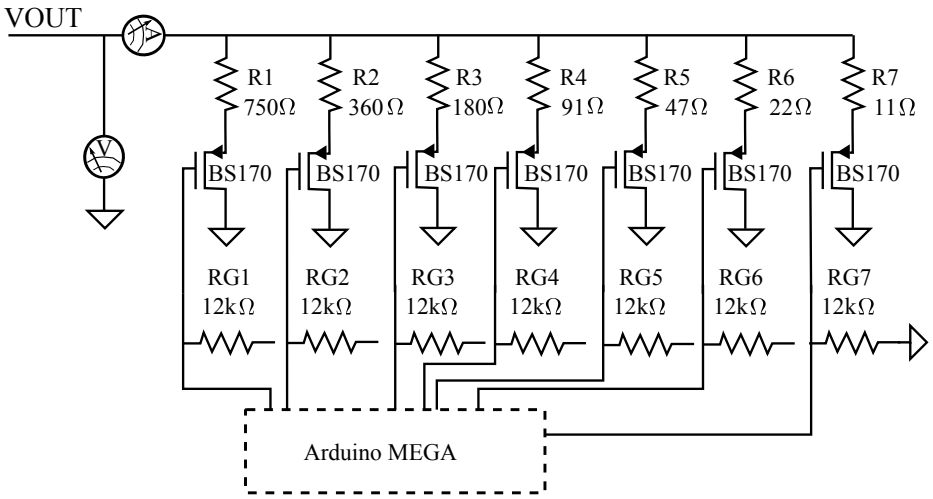


Figure 6.3. Schematic of the used dynamic load connected to the VOUT-pin of the TI BQ25570 chip.

6.2 Measurement Results

The system's average DC output voltage and the system's efficiency are measured, and the results are presented in this section. For the presented measurements, T_P is the pulse duration corresponding to a transmission of a frame and is set to 10 ms. T_i is the time between two consecutive bursts required to recharge the storage capacitor (CSTOR).

The efficiency, η_{system}, of the complete harvesting system is expressed as:

$$\eta_{\text{system}} = \frac{P_{\text{out}}}{P_{\text{RF}}}, \tag{6.3}$$

where P_{RF} is the RF power provided by the signal generator, and P_{out} is the average

DC output power. P_{out} is calculated as:

$$P_{out} = \frac{T_p}{T_p + T_i} P_L. \tag{6.4}$$

The system output voltage is VOUT=1.8 V and 55 mW (P_L = 55mW) of power is drawn by the dynamic load. A commercially available rectenna system [107] has been measured under the same conditions.
Figure 6.4 shows the average DC output power as a function of RF input power level for the commercially available rectenna system and for the presented system. The figure shows that the DC output power of the presented system outperforms that of the commercially available rectenna system, especially at lower input power levels (≤ -6dBm). The presented system is able to generate 10 μW on average at an RF input power level of -12 dBm, while the commercially available rectenna system generates the same output power at an input power level of -6 dBm, which results in 6 dB sensitivity improvement for this average DC output power level.

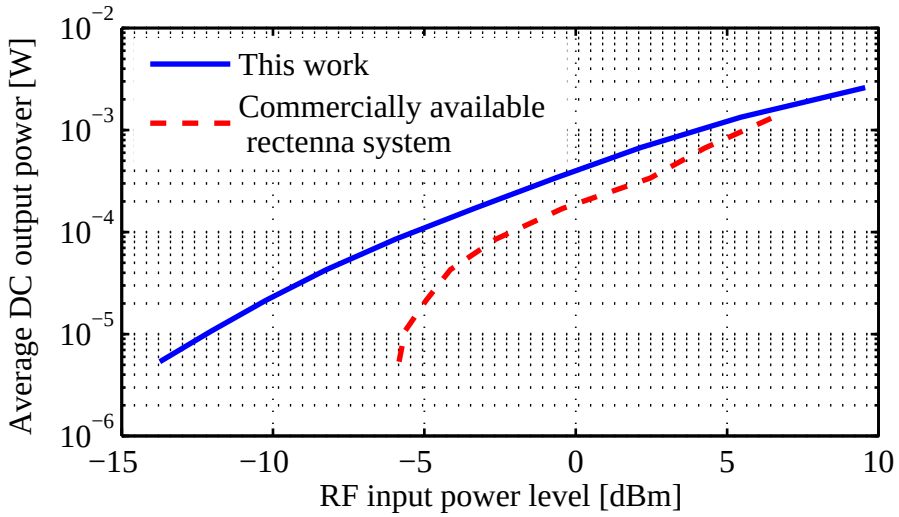


Figure 6.4. Average DC output power as a function of RF input power level using the harvesting system developed in this thesis and the commercially available rectenna system. PL=55mW, T_p = 10ms, T_i varies.

The system's RF to DC conversion efficiency is calculated using Eqs.(6.3) and (6.4) and the results for both systems are plotted in Fig. 6.5. The figure shows that the power conversion efficiency of the presented system is higher than that of the commercially available system in the shown RF power level band. The efficiency of the presented system reaches 23 % at an input power level of -10 dBm, while at the same input power level the commercially available system fails to operate.
At an input power level of 0 dBm, the power conversion efficiency of the commercially

available system is 19 %, while the power conversion efficiency of the presented system reaches 40 % resulting in more than 20 % improvement in power conversion efficiency.

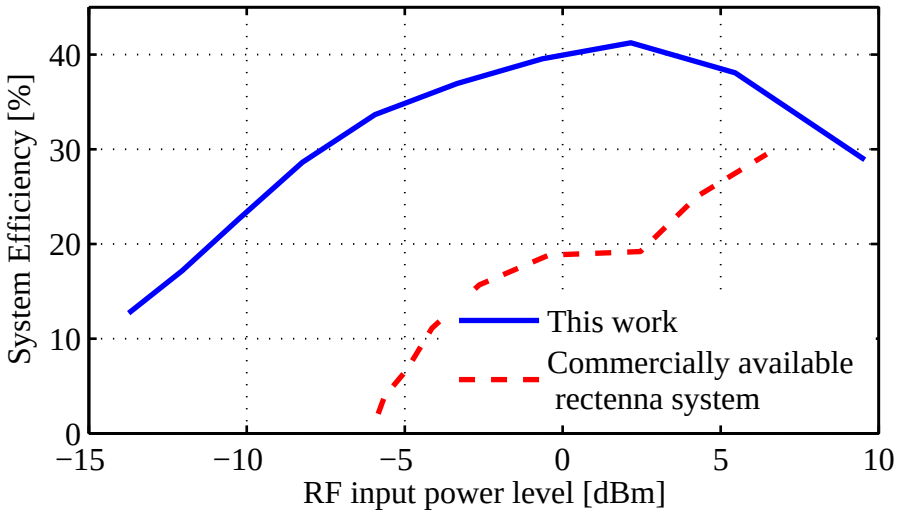


Figure 6.5. System's RF to DC conversion efficiency as a function of RF input power level using the harvesting system developed in this thesis and the commercially available rectenna system. PL=55mW, T_p = 10ms, T_i varies.

Using the measured system efficiency (solid curve in Fig. 6.5), the average DC output power is calculated as a function of rectenna's distance from a 2 W Effective Radiated Power (ERP) source at 868 MHz. Figure 6.6 shows the calculated average DC output power as a function of propagation distance when using a 1 dBi and a 6 dBi receiving antenna.

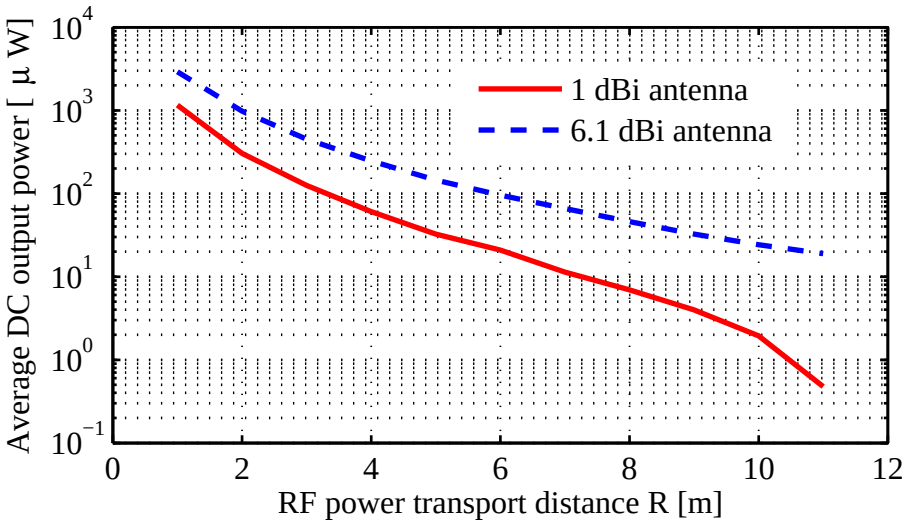


Figure 6.6. Average DC output power as a function of RF power transport distance (R) using a 1 dBi and a 6 dBi receiving antenna.

Figure 6.6 indicates that for an average DC output power of 20 μW, the RF power transport distance increases from 6 meters to 10.5 meters when using a 6.1 dBi gain antenna instead of 1 dBi gain antenna. An average DC output power of 20 μW, having a P_L of 55 mW and a burst time $T_p = 10$ms results in an off-time of $T_i = 27$ seconds.

As a practical application, the commercially available humidity and temperature sensor, Cresta WX688 [108] is used. The Cresta sensor is shown in Fig. 6.7. The sensor uses a 433 MHz transmitter to send data to the main station. The (two AA 1.5 V) batteries of the sensor are removed and the power lines are extended as shown in Fig. 6.7, and connected to the VOUT pin of the TI BQ25570 chip (see Fig. 6.2).

Figure 6.7. Cresta humidity and temperature sensor.

To provide 3.0 V at the VOUT-pin of the TI BQ25570, the resistor network is modified where R_{OUT2} and R_{OUT1} (see Fig. 6.2) are set to 7.5 MΩ and 5.1 MΩ respectively. The sensor transmits a burst containing the temperature and the humidity information to the main station every 45 seconds ($T_i = 45$ seconds). The burst duration is 823 ms, ($T_p = 823$ ms). The voltage variation on the storage capacitor (CSTOR) is 844 mV (3.844 V to 3.0 V). The energy stored in the capacitor is calculated using Eq. (6.1) and is 2.89 mJ. This results in a 3.5 mW power drawn by the sensor and an average DC output power, using Eq. (6.4), of 62.9μW.

The sensor operated at different RF input power levels. The minimum RF input power level that operates the sensor was -5 dBm[2]. It should be mentioned that the sensor

[2]The system efficiency is 35 % at -5dBm for drawing 55 mW during 10 ms (see Fig. 6.5). The system efficiency is expected to decrease if 3.5 mW are drawn during 823 ms. Since the minimum RF input power level that operated the sensor was -5 dBm (316μW), and the average DC power drawn by the sensor is 62.9μW, this results in a system efficiency of 20 % for powering the Cresta humidity and temperature sensor, which is consistent with our expectations.

is not developed for RF power transport applications, but the sensor was used in this experiment as a proof of concept.

6.3 Conclusion

In this chapter, a complete rectenna system was presented. The system encompasses a matching network, a voltage doubling rectifier, a commercially available power management circuit, a 1000 μF storage capacitor and a dynamic load. The average DC output power and the system's RF to DC conversion efficiency are measured, and compared to those measured for a commercially available rectenna system operated under the same conditions. It has been shown that the commercially available rectenna system fails to operate at low power levels (≤ -6dBm) while the presented system can operate at an RF input power level as low as -13 dBm. This means that for example, 55 mW of DC power can be drawn for 10 ms every 27.5 seconds on a distance of about 10 m from a 2 W ERP source at 868 MHz using a 6.1 dBi receiving antenna.
At an input power level of 0 dBm, the power conversion efficiency of the commercially available system does not exceed 19 %, while the power conversion efficiency of the presented system reaches 40 %, resulting in more than 20 % improved power conversion efficiency.

Chapter seven

A Frequency Selective Surface for RF Energy Harvesting

In this chapter, a novel implementation and application of Frequency Selective Surfaces (FSS) is presented. The novel FSS design can harvest power from arbitrarily polarized incident electromagnetic waves. The design does not involve a matching network, which results in a simple (polarization-independent) power harvester. An equivalent circuit model to predict the transmission and reflection characteristics of a gridded square loop FSS loaded with diodes is employed. The design method is validated for different incident angles, through full-wave analysis and measurements. To that end, the addition of lumped elements (R,L,C) in a metallic conductive grid of a non-harvesting FSS is investigated. Waveguide simulator measurements and full-wave simulations are employed to validate the derived analytical equations. A 3×3 and a 5×5 RF harvesting FSS have been designed, fabricated and measured. An RF to DC conversion efficiency of 25 % for the 3×3 RF harvester and 16% for the 5×5 RF harvester are measured at an incident power density of $3.2\,\mu\mathrm{W/cm^2}$.

7.1 Introduction

In the previous chapters the developed rectennas are designed to harvest linearly polarized waves. In this chapter a practical and efficient way to harvest RF waves that may be polarized in any direction is proposed. The technique can be used to produce RF power harvesting glass, or RF power harvesting wall paper. Figure 7.1 shows a photograph of two RF power harvesting glass prototypes using two different Frequency Selective Surfaces (FSS): the gridded square FSS and the gridded cross FSS. In the final design, transparent conductive film [109, 110] could be implemented on glass for the realization of optically transparent harvesting grids.

Figure 7.1. RF power harvesting glass using frequency selective surfaces.

The main advantages of the suggested RF harvester in comparison to conventional rectenna array configurations, (i.e. wherein each element is connected to a rectifier and wherein the DC-signals are combined) are as follows.

- In a conventional rectenna array configuration, RF combining networks or/and DC combining circuits are needed e.g. [111]. In the FSS-based rectenna, the combining circuits are absent. The DC output voltage is collected using only two DC connections to the structure, which simplifies the design of the RF harvester.
- The matching networks used in conventional rectennas and rectenna arrays are in general narrow-band and therefore enable power harvesting in most cases in a single frequency band only e.g. [112]. In the FSS harvesters presented in this chapter, impedance matching networks are not present. They form an integral part of the FSS structure, which eliminates the losses that would be introduced

by a separate matching network. In addition, the suggested design allows power harvesting from a dual frequency band, more specifically: the FSS transmission and reflection frequency bands.

- To maximize the harvested power, polarization matching between the transmitter and the receiver is crucial. Consequently, dual linearly, or circularly polarized rectennas are designed e.g. [113]. In this chapter, a polarization independent FSS is presented, which allows to harvest power from arbitrary polarized waves by decomposing the polarization of the incident wave into two orthogonal linear polarizations and rectifying them independently.

- Array rectennas (i.e. array antennas connected to a diode) exhibit a relatively high directivity and thus need to be positioned in the direction of the source for an optimal power transfer. The suggested FSS harvester that is a rectenna array (i.e. combination of radiators, each one connected to a rectifier) is less sensitive to angles of incidence and can harvest RF power efficiently at angles away from broadside.

The FSS rectenna discussed in this chapter consists of a metallic grid with inner square loops of conductors disposed in a periodic fashion as shown in Fig. 7.2(a).

The grid is loaded with Schottky diodes for current rectification. After rectification, the DC power is collected at the top-left and bottom-right corners as shown in Fig. 7.2(a). The physical principles of operation were first discussed in [114] and are shown in Figs. 7.2(b) and 7.2(c). For our FSS configuration, the labels U (up), D (down), R (Right) and L (Left) indicate the direction of the induced current in the grid. To start with, a vertically polarized incident wave $\overrightarrow{E_V}$ is considered, see Fig.7.2(b). The induced current will flow from the diodes d_{V1} toward diodes d_{V2} in the U-direction. Diodes d_{H1} and d_{H2} are unbiased, and consequently they will prevent the current from flowing in the R-direction. Similarly, when a horizontally polarized incident wave $\overrightarrow{E_H}$ hits the FSS surface, see Fig.7.2(c), this time diodes d_{H1} and d_{H2} draw the current in the L-direction from d_{H1} toward d_{H2}. Diodes d_{V1} and d_{V2} are unbiased, and consequently they will prevent the current from flowing in the D-direction. When an elliptically polarized wave hits the FSS surface, the vertical and the horizontal components of the electric field will drive the induced current to flow in the U and L-directions. Consequently the induced current can be collected at the top left corner by using the DC collection lines as shown in Fig.7.2(a). With this configuration the grid acts as a half-wave rectifier, making the FSS capable of rectifying the signal from arbitrarily polarized waves.

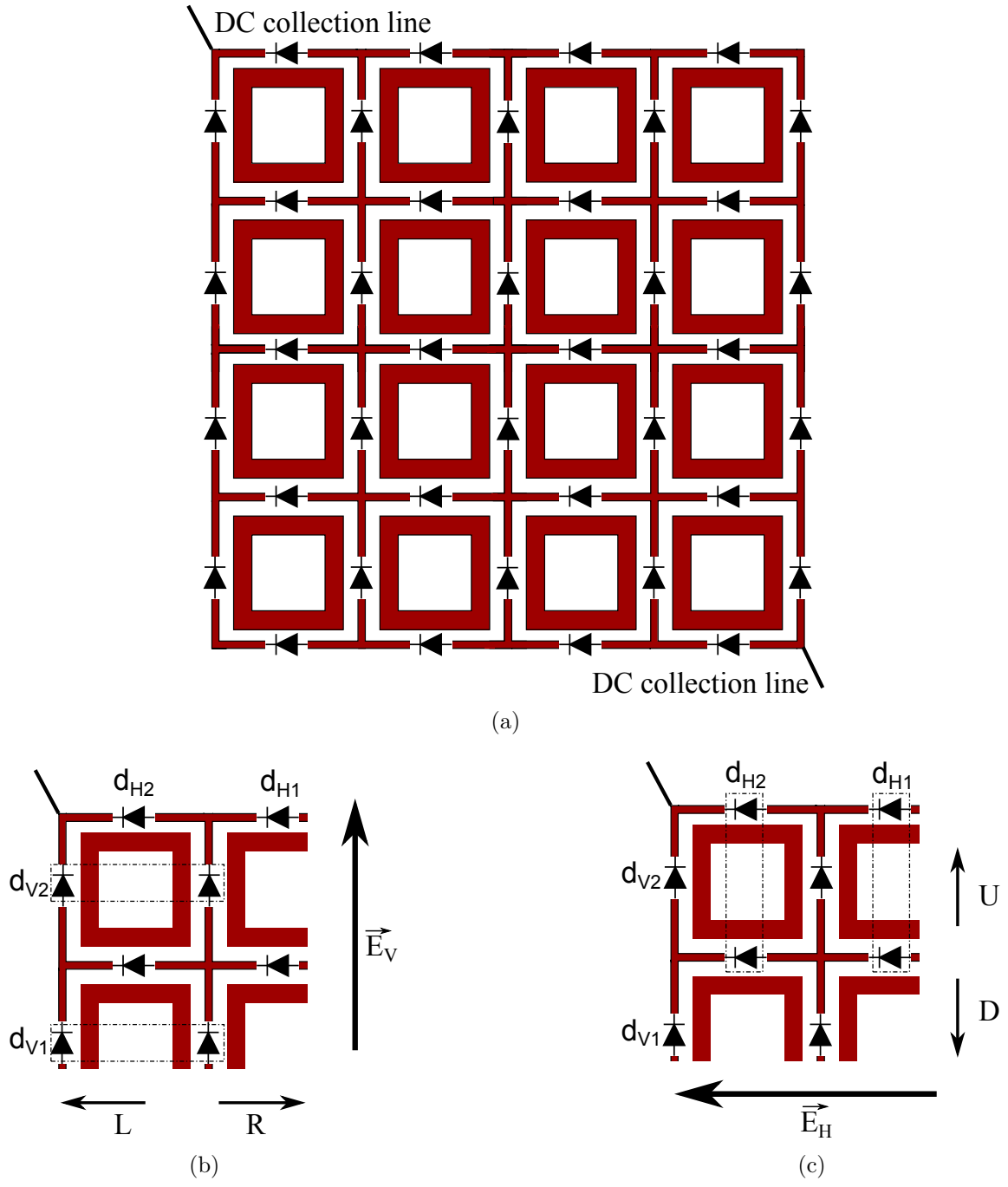


Figure 7.2. Suggested RF FSS harvester (a) and its physical principal of operation for an incident vertical electric Field (E_V), (b) and an incident horizontal electric Field (E_H), (c).

7.2 Square Loop FSS

The square loop FSS has been reported in the literature by several authors [115, 116] and is shown in Fig. 7.3(a).

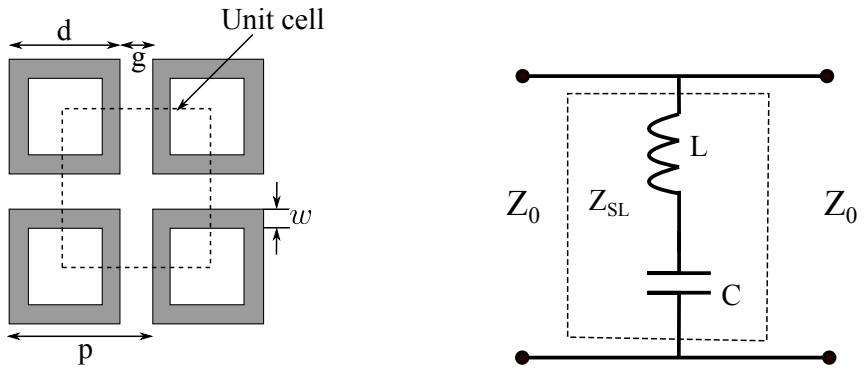


Figure 7.3. Square Loop (SL) frequency selective surface (FSS) configuration (a) and its unit-cell equivalent electric circuit model (b).

It consists of a periodic array of identical conducting strip square loops. g is the gap between neighboring elements, w is the width of the loop strip, d is the length of the square side and $p = (d + g)$ is the periodicity of the unit cell. The unit cell is the structure that is periodically repeated in the two orthogonal planar directions. Several choices can be made, the results are identical. The equivalent circuit model of the unit cell is shown in Fig. 7.3(b) and is a series LC circuit, shunt-positioned in a segment of transmission line carrying a plane incident wave, where L represents the inductance due to the vertical (horizontal) metal conductor and C is the capacitance between the horizontal (vertical) conductors.

The impedance of the unit cell (dashed box in Fig. 7.3(a)), is given by:

$$Z_{fss} = j\left(X_L - \frac{1}{B_C}\right), \tag{7.1}$$

where X_L and B_C are derived as presented in [117]:

$$X_L = Z_0\left(\frac{d}{p}\right) F\left(p,\, 2w,\, \lambda\right), \tag{7.2}$$

$$B_C = \frac{4}{Z_0}\left(\frac{d}{p}\right) F\left(p,\, g,\, \lambda\right). \tag{7.3}$$

Z_0 is the characteristic impedance of free space, $Z_0 = \sqrt{\frac{\mu_0}{\epsilon_0}}$ and λ is the free-space wavelength. The function $F(p, g, \lambda)$ is given in [118] and [119]:

$$F_{TE}(p, w, \lambda) = \Upsilon \left[\ln \csc \left(\frac{\pi w}{2p} \right) + G(p, w, \lambda, \theta) \right], \tag{7.4}$$

$$F_{TM}(p, w, \lambda) = \Phi \left[\ln \csc \left(\frac{\pi w}{2p} \right) + G(p, w, \lambda, \varphi) \right], \tag{7.5}$$

$$F_{TE}(p, g, \lambda) = \Psi \left[\ln \csc \left(\frac{\pi g}{2p} \right) + G(p, g, \lambda, \theta) \right], \tag{7.6}$$

$$F_{TM}(p, g, \lambda) = \Omega \left[\ln \csc \left(\frac{\pi g}{2p} \right) + G(p, g, \lambda, \varphi) \right], \tag{7.7}$$

where $\Upsilon = (p \cos\theta)/\lambda$, $\Phi = (p \sec\varphi)/\lambda$, $\Psi = (p \sec\theta)/\lambda$ and $\Omega = (p \cos\varphi)/\lambda$. Herein, θ and φ are the angles of incidence of a TE and TM polarized wave, respectively. The FSS is supposed to be positioned in the xy-plane of a rectangular xyz coordinate system. G is the first-order correction term for both TE-wave and TM-wave incidence, and has been given in [118] and [120]. The correction factor G is presented in Appendix B.
To account for the influence of the supporting dielectric layer, the results derived in [121] and [122] are used and presented in Sec. 7.2.2. The absolute value in dB of the transmission and reflection coefficients is derived from the normalized impedance of the unit cell [120]:

$$|T|_{dB} = 20 \log_{10} \left[\frac{2Z_{fss}}{1 + 2Z_{fss}} \right], \tag{7.8}$$

$$|R|_{dB} = 20 \log_{10} \left[\frac{-1}{1 + 2Z_{fss}} \right]. \tag{7.9}$$

7.2.1 Model Verification

To demonstrate the accuracy of these Analytical Expressions (AE), the reflection and the transmission coefficients are calculated and compared to the results obtained by

the Finite Integration Technique (FIT) [72] as implemented in CST Microwave Studio. Table 7.1 summarizes the dimensions for different array configurations that are investigated here. t is the substrate thickness and ϵ_r is the substrate permittivity.

Table 7.1. Analyzed Square Loop Array Configurations. Dimensions are explained in Fig. 7.3(a).

Square Loop Array #	**p** (mm)	**d** (mm)	**g** (mm)	**w** (mm)	**t** (mm)	$\epsilon_\mathbf{r}$
1	10.0	8.75	1.25	0.625	0.0	1.0
2	36	31.0	5	3.0	0.25	2.2
3	22.0	21.0	1.0	3.0	see Sec. 7.2.2	2.2
4	32.25	31.0	1.25	3.0	0.25	see Sec. 7.2.2

Figure 7.4 shows the magnitude (a) and the phase (b) of the transmission and the reflection coefficients for FSS #1 in Table 7.1 as a function of frequency calculated by the Analytical Expressions (AE) and by the Finite Integration Technique (FIT) using CST Microwave Studio, demonstrating the validity of the AE.

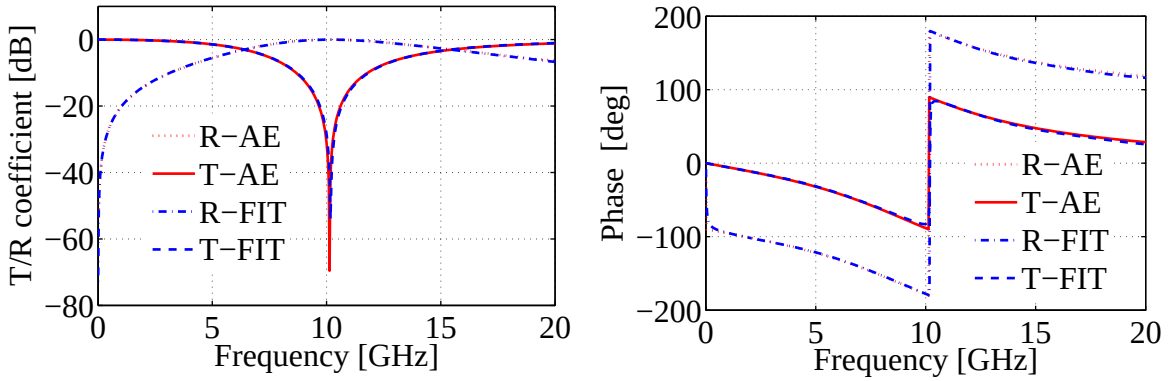


Figure 7.4. Magnitude (a) and phase (b) of reflection (R) and transmission (T) coefficients as a function frequency for the square loop Array #1 of Table 7.1 at normal incidence, $\theta = \varphi = 0°$.

The frequency response of the FSS is affected by the polarization of the incident plane wave (TE, TM) and by its angle of incidence (θ, φ). The main advantage of the presented equivalent circuit model is that it offers the capability of deriving expressions for predicting the behavior of the reflection/transmission coefficients for different polarizations and for an oblique angle of incidence, up to an angle of incidence of 45 degrees [119]. To demonstrate that the equivalent circuit model can accurately predict

the reflection and the transmission coefficients for different angles of incidence and for different polarizations, the square loop array # 2 presented in Table 7.1 is investigated. The frequency response is calculated with the aid of the presented analytical expressions and is compared to the results obtained by CST Mcirowave studio. Figure 7.5 shows the magnitude (a) and phase (b) of the reflection (R) and transmission (T) coefficients as a function frequency for the square loop Array #2 of Table 7.1, for a TE incidence with $\theta = 15°$, $\varphi = 0°$. The results for the same FSS are also calculated for a TM wave incidence with $\theta = 0°$, $\varphi = 15°$ and are plotted in Fig. 7.6, again demonstrating the validity of the AE.

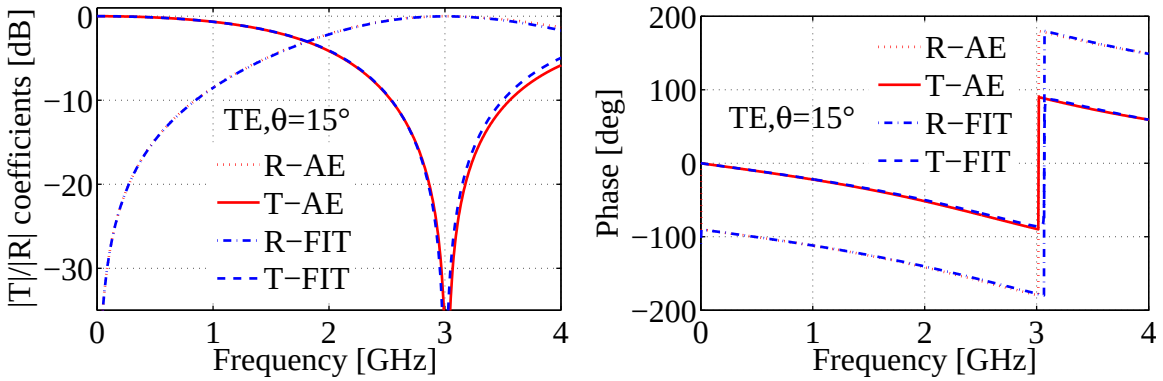


Figure 7.5. Magnitude (a) and phase (b) of reflection (R) and transmission (T) coefficients as a function frequency for the square loop Array #2 of Table 7.1, $\theta = 15°$, $\varphi = 0°$.

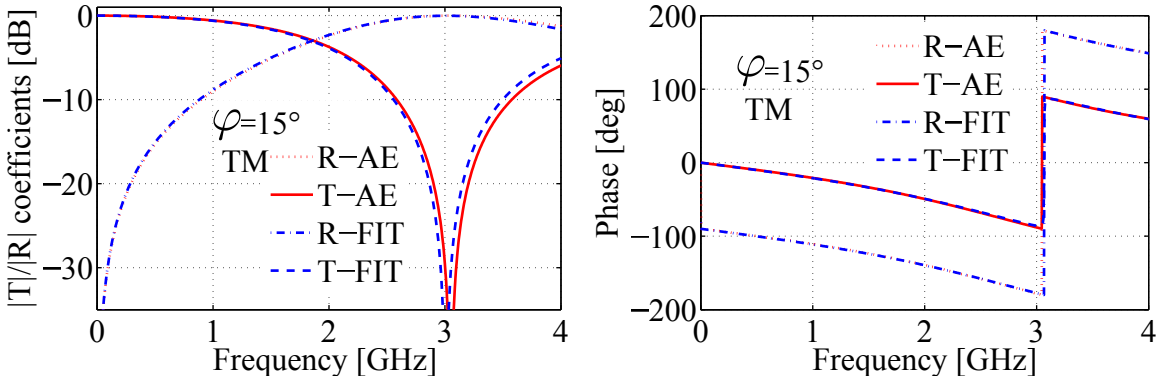
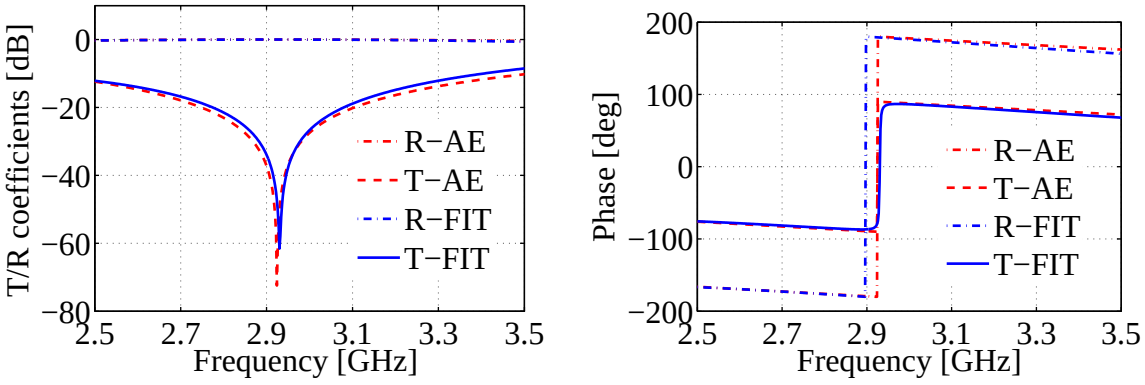


Figure 7.6. Magnitude (a) and phase (b) of reflection (R) and transmission (T) coefficients as a function frequency for the square loop Array #2 of Table 7.1, $\theta = 0°$, $\varphi = 15°$.

We conclude with a final example, wherein we look at the FSS #2 in Table 7.1 for an

angle of incidence, $\theta = 45°$, and $\varphi = 0°$. The reflection and the transmission coefficients are calculated using the equivalent circuit model and are compared with the results obtained by the FIT with the aid of CST Microwave Studio. Figure 7.7 shows the magnitude (a) and phase (b) of reflection (R) and transmission (T) coefficients as a function frequency for the square loop Array #2 of Table 7.1 at $\theta = 45°$, and $\varphi = 0°$.



(a) (b)

Figure 7.7. Magnitude (a) and phase (b) of reflection (R) and transmission (T) coefficients as a function frequency for the square loop Array #2 of Table 7.1 at $\theta = 45°$, and $\varphi = 0°$.

It is observed from Figs. 7.4-7.7 that the equivalent circuit model can predict the reflection and the transmission coefficients for a square loop FSS for different polarization and for different angles of incidence, with a frequency shift in the transmission and reflection coefficients less than 5%.

7.2.2 Effect of the Supporting Dielectric Substrate

The effect of the dielectric substrate that is carrying the array elements was discussed in [121–123]. For thin dielectric layers the value of the unit-cell capacitance, see Fig. 7.3(b), varies as a function of dielectric thickness t. A value of the effective permittivity ε_{eff}, which will be used instead of the substrate permittivity ε_r, is calculated via an iterative procedure for thin dielectric substrates in [121]. The expression depends on both ϵ_r and t and the resulting effective permittivity can be included in the equivalent circuit model by multiplying the susceptance B_C (Eq. (7.3)) by this effective permittivity. The expression of the effective permittivity (ε_{eff}) is obtained as:

$$\varepsilon_{eff} = \varepsilon_r + (\varepsilon_r - 1) \cdot \left[\frac{-1}{e^{x \cdot N}}\right], \tag{7.10}$$

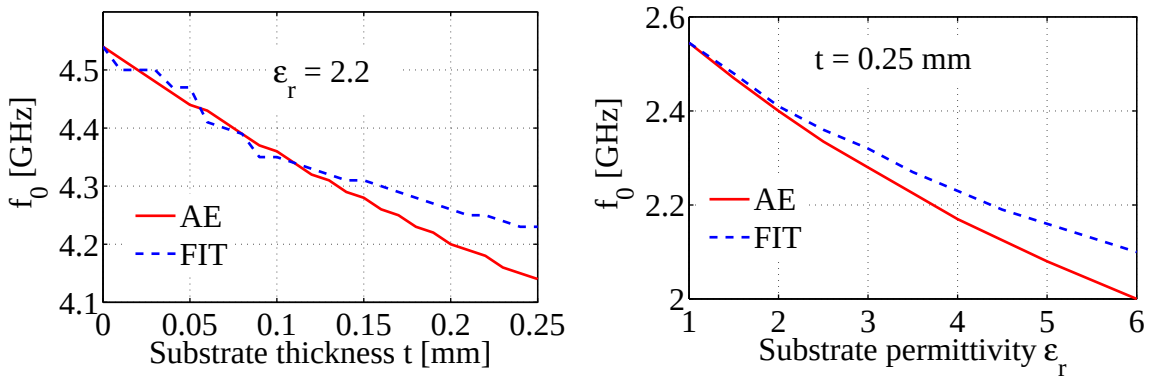
where

$$x = 10 \cdot t/p. \tag{7.11}$$

t is the thickness of the substrate, p is the periodicity and N is an exponential factor that depends on the geometry of the unit cell. For square loop arrays, N is approximately 1.8 [121]. For the square loop FSS the value of the susceptance is, therefore, modified as:

$$B_C = 4 \cdot \varepsilon_{eff} \cdot \frac{1}{Z_0} \left(\frac{d}{p}\right) F\left(p,\, g,\, \lambda\right). \tag{7.12}$$

To investigate the effect of the dielectric substrate, the resonant frequency (f_0), which is the transmission resonance frequency, is calculated by the equivalent circuit model and by FIT. The results are plotted as a function of thickness variation in Fig. 7.8(a) and as a function of substrate permittivity in Fig. 7.8(b).



(a) (b)

Figure 7.8. (a) Transmission resonance frequency as a function of substrate thickness t for the square loop Array #3 of Table 7.1. (b) Transmission resonant frequency as a function of substrate permittivity ε_r for the square loop Array #4 of Table 7.1. Both curves are calculated using the analytical equations (AE) and the Finite Integration Technique (FIT) at normal incidence. $\theta = \varphi = 0°$.

It is shown in the figures that the deviations between the presented analytical expressions and the Finite Integration Technique increases with increasing the substrate thickness (see Fig. 7.8(a)) and with increasing the substrate permittivity (see Fig. 7.8(b)). The relative difference error in the resonant frequency (f_0) between FIT and AE is calculated for a substrate thickness of 0.25 mm and $\varepsilon_r = 2.2$ (see Fig. 7.8(a)) and for a substrate permittivity of $\varepsilon_r = 6$ and a substrate thickness of 0.25 mm (see Fig. 7.8(b)) as are 2.12 % and 4.7 %, respectively.

Since the relative difference in the the resonant frequency (f_0) between FIT and AE less than 5%, then the results of the analytical expressions are considered to be accurate for a substrate thickness less than 0.25 mm and substrate permittivity ε_r between $1 \leq \varepsilon_r \leq 6$.

Since the equivalent circuit model is most accurate for a thin and low permittivity sub-

strate, for the fabricated RF power harvester a substrate thickness of 0.125 mm with a relative permittivity equal to 3.0 is chosen.

Having established the validity of the analytical expressions, we now proceed with the investigation of a gridded square-loop FSS. The addition of the grid to the loops is crucial for RF power harvesting since the diodes will be integrated in the grid.

7.3 Gridded Square Loop FSS

The frequency selective surface that will be used to harvest RF power is the gridded square-loop FSS (see Fig. 7.2(a)). A grid is added to the square-loop FSS introduced in the previous section. The main advantage of this structure is that, when it is loaded with rectifiers, the grid will direct the DC current to the collection point. Moreover, it provides additional physical parameters (width of the grid and grid-loop separation) that can be tuned to control the response of the FSS.

7.3.1 Equivalent Circuit Model

The equivalent circuit model introduced in the previous section is extended for the grid to predict the transmission and the reflection band behaviour for a Gridded Square-Loop (GSL) FSS. The FSS unit cell of the GSL-FSS is shown in Fig. 7.9(a).

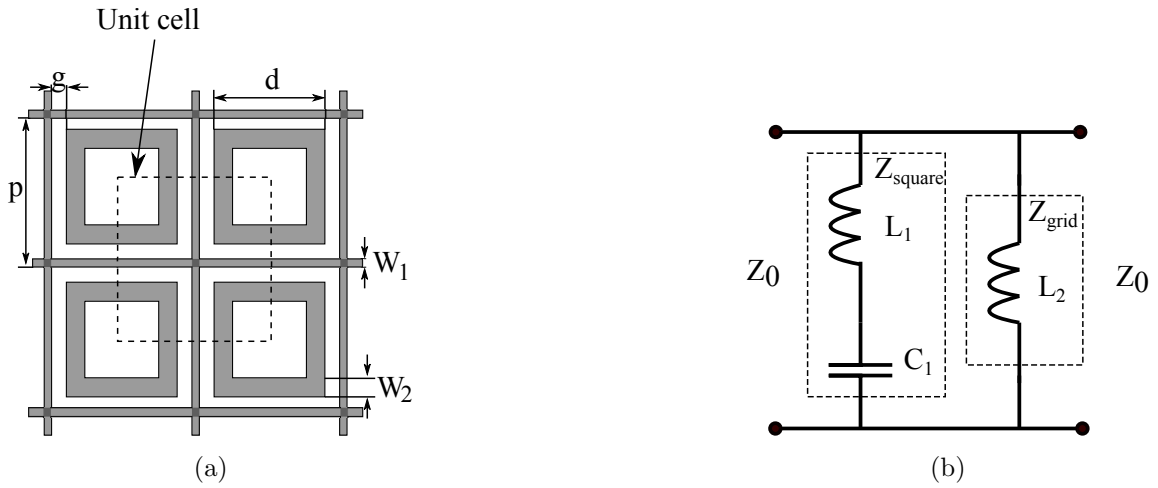


Figure 7.9. Unloaded (i.e. without diodes) gridded square loop configuration (a) and its unit cell equivalent circuit model (b).

In Fig. 7.9(a), p is the periodicity, g is the gap width between the grid and the square loop, W_1 is the strip width of the grid and W_2 and d are the strip width and the

length of the square loop, respectively. The equivalent electric circuit model introduced in [119, 124] is shown in Fig. 7.9(b). Z_{fss} is the equivalent FSS unit cell impedance and is derived by considering two impedances in parallel:

$$Z_{fss} = \frac{Z_{grid}\ Z_{square}}{Z_{grid} + Z_{square}}. \tag{7.13}$$

Z_{square} is the square-loop impedance and is given by:

$$Z_{square} = j\left(X_1 - \frac{1}{B_1}\right), \tag{7.14}$$

Z_{grid} is the grid impedance given by:

$$Z_{grid} = jX_2, \tag{7.15}$$

where [119]

$$X_1 = 2Z_0\left(\frac{X_2X_3}{X_2 + X_3}\right), \tag{7.16}$$

$$X_2 = Z_0F_{TE,TM}\left(p, W_1, \lambda\right), \tag{7.17}$$

$$X_3 = Z_0\frac{d}{p}F_{TE,TM}\left(p, 2W_2, \lambda\right), \tag{7.18}$$

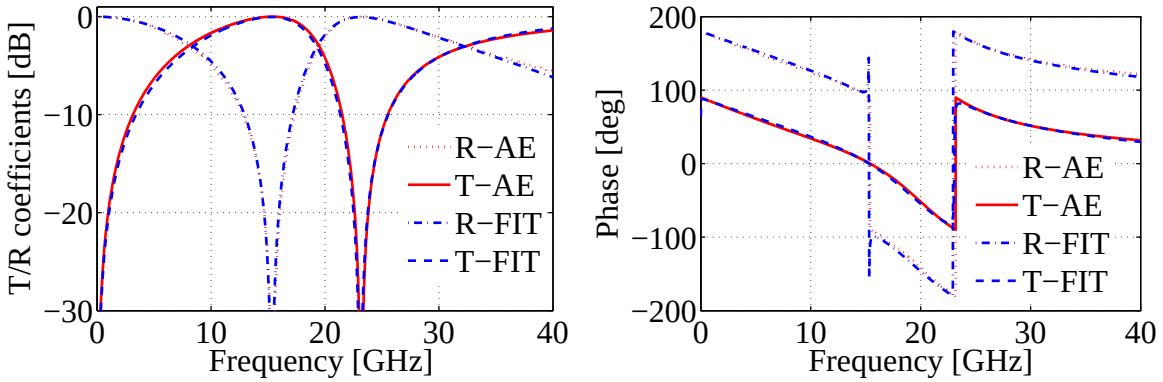
$$B_1 = \frac{2}{Z_0}.\epsilon_{eff}\ .\ \left(\frac{d}{p}\right)F_{TE,TM}\left(p, g, \lambda\right). \tag{7.19}$$

$F_{TE,TM}\left(p, g, \lambda\right)$ is a function of the incident angles θ and φ and depends on whether a TE incidence or TM incidence or if a combination of the two occurs. $F_{TE,TM}\left(p, g, \lambda\right)$ is calculated by using Eq. (7.4) to Eq. (7.7). Again, The FSS is supposed to be positioned in the xy-plane of a rectangular xyz coordinate system.

7.3.2 Model Verification and Waveguide Simulator Measurements

In this section, the equivalent circuit model presented in the previous section is tested by full-wave simulations and by measurements. The reflection and the transmission coefficients of a Gridded Square-Loop (GSL) FSS possessing the following dimensions: $p = 5.05\,\text{mm}$, $d = 3.7\,\text{mm}$, $g = 0.6\,\text{mm}$, $W_1 = 0.15\,\text{mm}$, $W_2 = 0.15\,\text{mm}$, are calculated by using the equivalent circuit model and the Finite Integration Technique. The GSL-FSS is positioned on a 0.027 mm thick substrate ($t = 0.027\,\text{mm}$) having a relative

permittivity of $\epsilon_r = 3.0$. Fig. 7.10 shows the magnitude (a) and the phase (b) of the reflection (R) and transmission (T) coefficients as a function frequency for the gridded square-loop array calculated by the FIT and by the Analytical Expressions (AE) for a TE incidence with $\theta = 0°$, and $\varphi = 0°$.



(a) (b)

Figure 7.10. Magnitude (a) and phase (b) of reflection (R) and transmission (T) coefficients as a function frequency for a gridded square loop array with $p = 5.05\,\text{mm}$, $d = 3.7\,\text{mm}$, $g = 0.6\,\text{mm}$, $W_1 = 0.15\,\text{mm}$, $W_2 = 0.15\,\text{mm}$, $t = 0.027\,\text{mm}$, $\epsilon_r = 3.0$. TE incidence with $\theta = 0°$, and $\varphi = 0°$.

To validate the accuracy of the analytical expressions with measurements, the waveguide-simulator technique is used [125]. The reflection and transmission coefficients versus frequency for a GSL-FSS unit cell with $d = 31\,\text{mm}$, $g = 2\,\text{mm}$, $W_1 = 1\,\text{mm}$, $W_2 = 4\,\text{mm}$ are calculated and compared to waveguide simulator measurements. The chosen dimensions of the FSS unit cell are dictated by the waveguide aperture used for measurements, see Fig. 7.11.

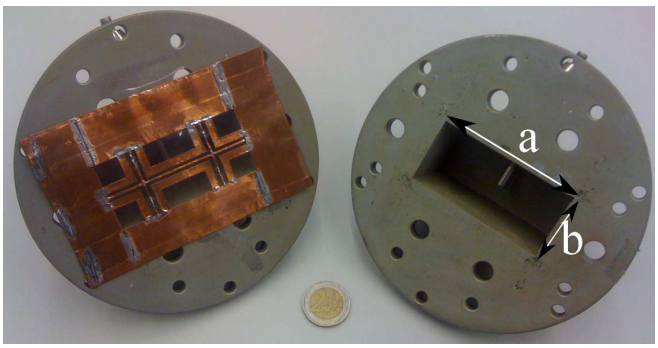


Figure 7.11. Waveguide used for measurements and the fabricated two unit cells matching the waveguide aperture.

The rectangular waveguide (SIVERS-LAB PM 7325 S-04) has side lengths $a = 71.5$ mm and $b = 34$ mm. The waveguide aperture contains two GSL-FSS unit cells. The lower and the higher cut-off frequencies of the waveguide are 2.5 GHz and 3.9 GHz, respectively. The FSS unit cell was designed to create a reflection band around 3.2 GHz.

The waveguide simulator reproduces the behaviour of an infinite array by exploiting the symmetry of a unit cell. The walls of the waveguide act like periodic boundary conditions around the elements [125]. The decomposition of the TE_{10} mode into two plane waves in a rectangular waveguide having the widest dimension a as introduced in [126] is shown in Fig. 7.12.

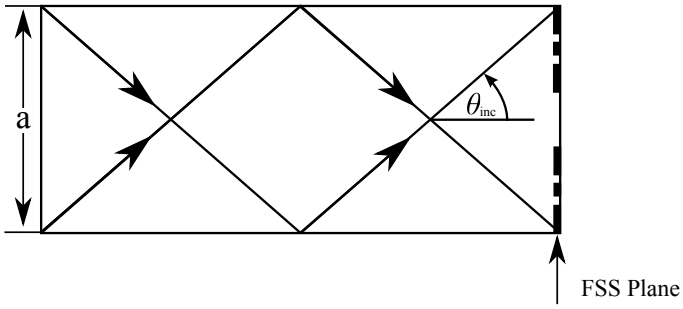


Figure 7.12. Decomposition of the TE_{10} mode into two plane waves in a rectangular waveguide having the widest dimension a as introduced in [126].

The incident angle θ_{inc} on the FSS plane, of the fundamental TE_{10} mode, traveling inside the waveguide, depends on: the operating frequency f, the widest dimension of the rectangular waveguide a, and the free-space speed of light c. The angle of incidence is calculated from the following expression [127]:

$$\theta_{inc}\,[\text{degrees}] = \frac{180}{\pi}\cos^{-1}\left(\frac{\sqrt{\left(\frac{2\pi \mathrm{f}}{\mathrm{c}}\right)^2 - \left(\frac{\pi}{\mathrm{a}}\right)^2}}{\frac{2\pi \mathrm{f}}{\mathrm{c}}}\right). \tag{7.20}$$

From Eq. (7.20), at a frequency of 3.2 GHz, using $a = 71.5$mm, the incident angle on the FSS surface is $\theta_{inc} = 41°$. It should be mentioned that, in order to fit the two fabricated unit cells in the waveguide aperture, a slight asymmetry in the fabricated unit cell is introduced. Fig. 7.13 shows the waveguide simulator setup and specifies this asymmetry.

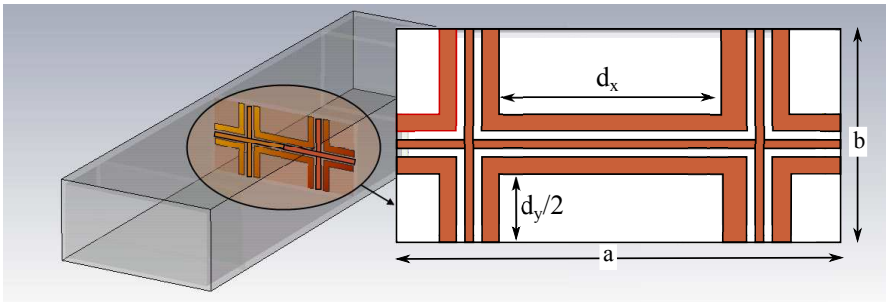


Figure 7.13. Waveguide simulator setup of two unit cells of a GSL-FSS array. $d_x = 31$ mm, $d_y = 30$ mm.

Referring to Fig. 7.13, the dimensions chosen for the fabricated GSL-FSS are $d_y = 30$ mm and $d_x = 31$ mm. With the introduction of the small asymmetry in the unit cell, the elements are now matching the waveguide aperture.

Figure 7.14 shows the simulated and measured magnitudes of the transmission (a) and reflection (b) coefficients as a function of frequency at an angle of incidence of $\theta_{inc} = 41°$.

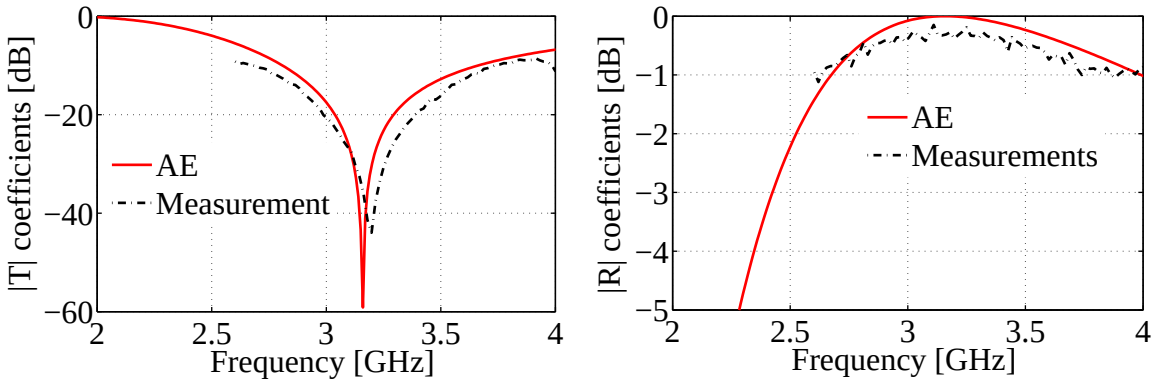


Figure 7.14. Magnitude of the transmission (a) and reflection (b) coefficients versus frequency for a GSL-FSS unit-cell with $d = 31$ mm, $g = 2$ mm, $W_1 = 1$ mm, $W_2 = 4$ mm. TE with $\theta_{inc} = 41°$, as simulated with AE and measured using a waveguide simulator.

As depicted in the figure, a frequency offset exists between the measurement results and the simulated results. This frequency offset is believed to be due to the asymmetry introduced by the fabricated two unit cells in order to match the waveguide aperture. It should be mentioned that the two waveguide adapters (see Fig. 7.11) are tightly fastened to create a good galvanic contact between the unit cell edges and the waveguide walls. Referring to Fig. 7.14(a), the resonance frequency for the measured transmission

coefficient is $F_{T_{measured}} = 3.197$ GHz, while the resonance frequency for the transmission coefficients calculated by the equivalent circuit model is $F_{T_{AE}} = 3.16$ GHz, which results in an offset of 37 MHz. The relative difference frequency offset $\delta\,(\%)$ is calculated as follows:

$$\delta\,(\%) = 100\left(\frac{F_{T_{measured}} - F_{T_{AE}}}{F_{T_{measured}}}\right), \tag{7.21}$$

which is equal to 1.15 %. Since the relative difference frequency offset $\delta\,(\%)$ is less than 5 %, the equivalent circuit model is considered to be accurate and will be extended in the following section for the insertion of lumped elements.

7.4 Loaded Gridded Square-Loop FSS

The basic principle of creating a RF frequency selective surface harvester is to insert diodes in the grid. By resonating the FSS and rectifying the resonating RF currents, DC power is obtained. The harvested DC current is collected at the DC collection lines as shown in Fig. 7.2(a). In this section, the equivalent circuit model introduced in the previous section will be extended to account for the rectifiers (Schottky diodes) that are inserted to the grid.

7.4.1 Equivalent Circuit Model

Figure 7.15(a) shows a loaded unit-cell configuration. The reason for having two rectifiers for each polarization in the unit cell as indicated in the figure will be explained further on.

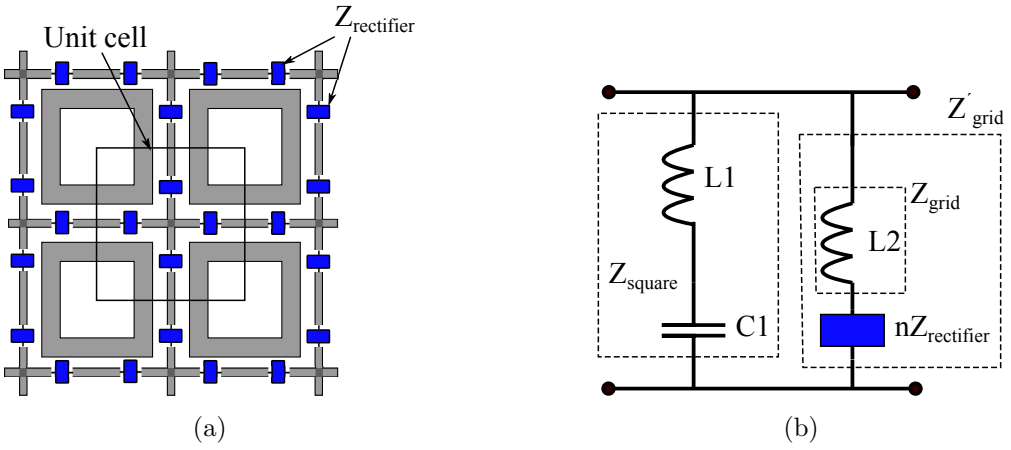


Figure 7.15. Loaded gridded square loop configuration (a) and its unit cell equivalent electric circuit model (b).

Since the rectifiers will be integrated to the grid, the equivalent impedance of the rectifier $Z_{rectifier}$ is placed in series with the grid impedance Z_{grid} in the equivalent unit-cell circuit, see Fig. 7.15(b). Combined, the grid impedance Z_{grid} and the rectifier's impedance $nZ_{rectifier}$ are denoted by Z'_{grid}. n is the number of lumped elements (Schottky diodes) in the unit cell for each polarization. For the unit cell shown in Fig. 7.15(a), two lumped elements for each polarization are shown, consequently n is set to 2. It should be mentioned that in the final design one rectifier per unit cell will be used for each polarization. The reason for having two rectifiers (Schottky diodes) has to do with the practical implementation of measuring a unit cell in a waveguide simulator. The unit-cell as implemented in the waveguide simulator is indicated in Fig. 7.15(a) and is chosen in this way to prevent the creation of a lumped element between the edges of the grid and the waveguide walls. Using a single rectifier would result in needing half a rectifier in the waveguide simulator at the edges of the grid of a unit cell as indicated in Fig. 7.16.

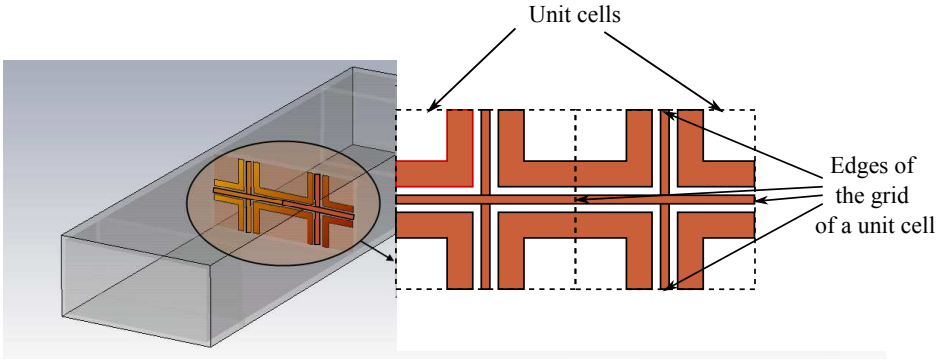


Figure 7.16. Two unit cells fitting the aperture of the waveguide simulator.

It has been demonstrated in Sec. 2.2 that the equivalent impedance of a diode, at an operating frequency below 3 GHz (see Figs. 2.2, 2.3 and 2.8) and at an input power level between -20 dBm and -5 dBm (see Fig. 2.9), may be modeled (linearized) as a series R-C circuit.
For a FSS loaded with an R-C series circuit, the FSS equivalent unit-cell impedance Z_{fss} is calculated from Eq.(7.13), where the grid impedance is extended to

$$Z'_{grid} = jX_2 + nZ_{rectifier} \tag{7.22}$$

$$= n\left[R_{rectifier} + j\left(\frac{X_2}{n} - \frac{1}{B_{rectifier}}\right)\right], \tag{7.23}$$

where $B_{rectifier} = \omega C_{rectifier}$, ω being the angular frequency. With expression 7.23 an FSS having a grid loaded with any combination of R, L and C can be analysed.

7.4.2 FSS Loaded with Capacitors

We start with a unit cell that is loaded with two capacitors in series with the grid inductance L2 (see Fig. 7.15(b)). Figures 7.17 and 7.18 show the transmission and reflection coefficients versus frequency respectively, for the thus loaded FSS for different capacitance values.

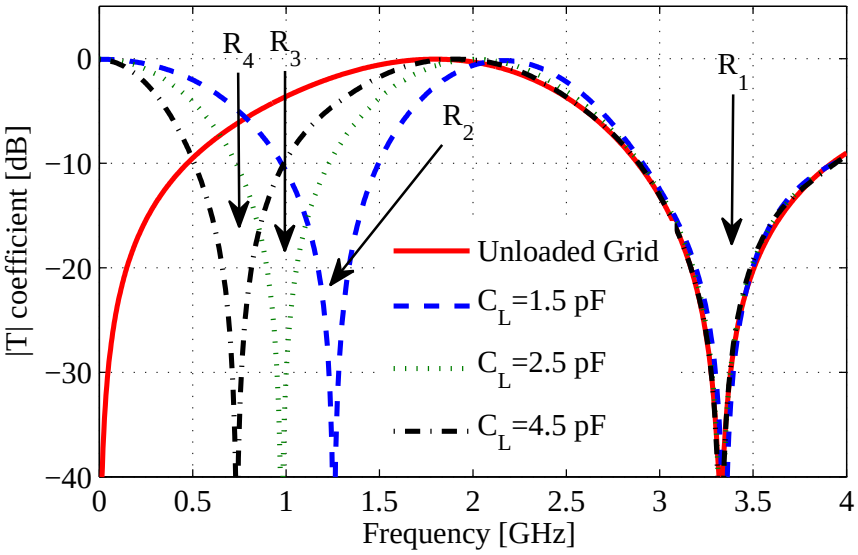


Figure 7.17. Magnitude of the transmission coefficients versus frequency for a gridded-square-loop unit-cell with $d = 31$ mm, $g = 2$ mm, $W1 = 1$mm, $W2 = 4.5$ mm. TE with $\theta_{inc} = 0°$.

In Fig. 7.17, the 'Unloaded Grid' label represents the gridded square loop investigated in Sec. 7.3. It is shown in the figure that for the unloaded grid (solid curve), only one reflection band is observed. The centre of the reflection band is indicated with R_1. This single reflection band is due to the capacitor C_1 placed in series with the inductor L_1 (see Fig. 7.15(b)), which creates a resonance at R_1. However, when the grid is loaded with a capacitor, for instance 1.5 pF (dashed curve) a second reflection band is generated around $R_2 = 1.25$ GHz. Consequently the loaded FSS is characterized by two reflection bands centered around R_1 and R_2. When the capacitance value is increased to 2.5 pF (dotted curve) and to 4.5 pF (dashed-dotted curve) two new reflection bands centered around R_3 and R_4 are created. Increasing the capacitance value thus results in a shifting of the additional reflection band to the lower frequencies. The created second resonance R_2 is due to the capacitor C_2 placed in series with the inductor L_2 (see Fig. 7.15(b)).

From the reflection coefficients shown in Fig. 7.18, it is observed that the unloaded grid (solid curve) resonates around 1.65 GHz (T_1). When a 1.5 pF capacitor is inserted

in the grid, the transmission band is shifted to the right to form a shifted-reflection band T_2. If the capacitance value is further increased to 2.5 pF and to 4.5 pF, the transmission band shifts back to the left toward T_3 (dotted curve) and T_4 (dashed-dotted curve). It is clear from the transmission and reflection bands that, when the capacitance value is further increased to a relatively high value, the loaded FSS will behave like the unloaded one, since high-valued capacitances will behave like a short circuit at the considered frequencies.

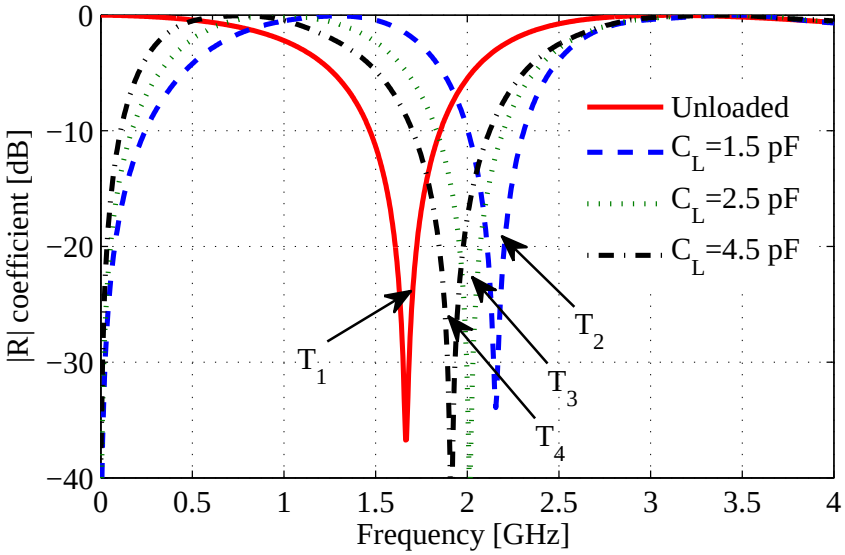
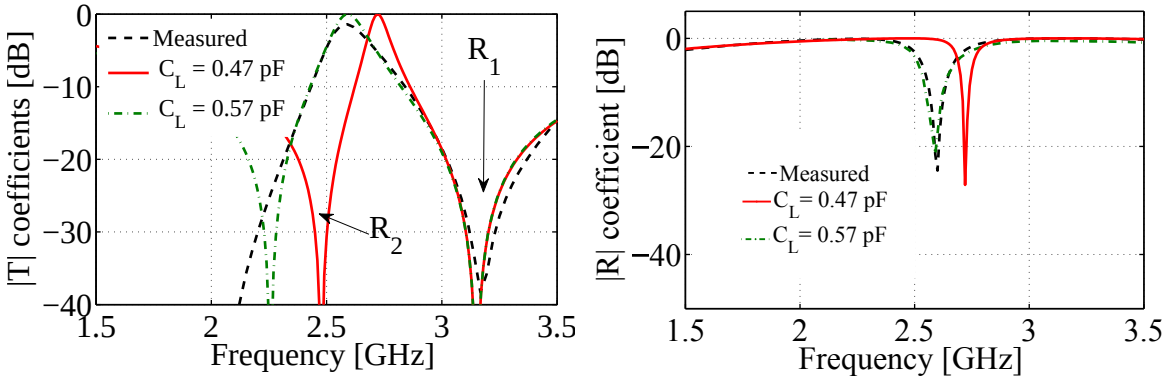


Figure 7.18. Magnitude of the reflection coefficients versus frequency for a gridded-square-loop unit-cell with $d = 31$ mm, $g = 2$ mm, $W1 = 1$mm, $W2 = 4.5$ mm. TE with $\theta_{inc} = 0°$.

A capacitively loaded GSL-FSS prototype is designed, simulated, manufactured as two slightly asymmetric unit-cells and measured. The gridded square loops are printed on a 0.125 mm thick substrate having a relative permittivity $\epsilon_r = 3.0$ and are loaded with four 0.47 pF capacitors in total positioned in the vertical direction. Fig. 7.19 shows the calculated and the measured transmission (Fig.7.19(a)) and reflection (Fig. 7.19(b)) coefficients as a function of frequency.

It is shown in Fig. 7.19(a) that the first resonance in the transmission coefficients around the centre of the reflection band R_1 is predicted by the analytical expressions. A slight deviation is observed between the results of the analytical expressions and the measured results, which is believed to be due to the asymmetry introduced in the manufactured unit cells to fit the waveguide aperture. However, for the second resonance centered around R_2, which is due to the capacitive load introduced in the grid, a large difference is observed in the centre frequency between the simulated and

the measured results for a capacitive load of 0.47 pF. A similar shift is observed in the transmission resonance shown in Fig. 7.19(b). This difference in the centre frequency of the measured responses is within the tolerance of the capacitor used for measurements (± 0.25pF) [128]. Within the tolerance value of the measured capacitor, it is shown in the same figures that for a capacitive load of 0.57 pF, the results of the analytical expressions match the measured results more closely.



(a) (b)

Figure 7.19. Magnitude of the transmission (a) and reflection (b) coefficients versus frequency for a capacitively loaded GSL-FSS simulated using the presented analytical expressions and measured using the waveguide simulator. $d = 31$ mm, $g = 2$ mm, $W_1 = 1$ mm, $W_2 = 4$ mm. TE with $\theta_{inc} = 41°$, $C_L = 0.47$ pF.

With the design strategy and the analytical expressions considered to be sufficiently validated, we now proceed to the design of a RF power harvesting FSS.

7.5 FSS Power Harvesting - Final Design

For the validation of the new concept, only harvesting power from a vertically polarized wave is considered since this simplifies the measurement setup. Consequently, the Schottky diodes are only placed in the vertical direction. At an operating frequency of 1GHz, the impedance of the diode is calculated for different P_{in} from the analytical expressions presented in Sec. 2.2. This impedance corresponds to an equivalent series circuit of a resistor and a capacitor. For an input power level P_{in} between $-14\,\text{dBm} \leq P_{in} \leq -5\,\text{dBm}$ the real part R of the input impedance and the capacitive part C are shown in Table 7.2.

Table 7.2. Real and capacitive parts of the HSMS-2820 Schottky diode as a function of input power level. The operating frequency is 1 GHz.

P_{in} [dBm]	R [Ω]	C [pF]
-14	5.66	1.17
-11	6.322	1.18
-8	10.85	1.19
-5	39.73	1.21

The rectifier's equivalent impedance is used in the equivalent circuit model introduced in the previous section to predict the reflection and the transmission behavior as a function of frequency. The reflection and the transmission bands will be used as the RF power harvesting bands. The parameters of the GSL-FSS are optimized for the new reflection band introduced by the loading (R_2 see Fig. 7.20(a)) to resonate around 1 GHz. Figure 7.20 shows the magnitude of the transmission (a) and reflection (b) coefficients as a function of frequency for the final GSL-FSS design for different RF input power levels. For every input power level the corresponding rectifier's impedance as stated in Table 7.2 is used. The equivalent circuit model is used for simulations.

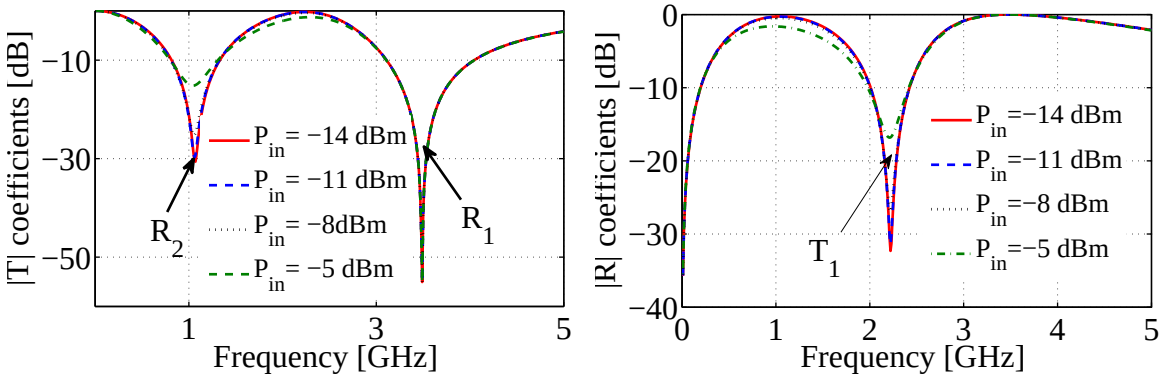


Figure 7.20. Magnitude of the transmission (a) and reflection (b) coefficients versus frequency for different P_{in}, TE incident wave $\theta_{inc} = 0°$. The optimized parameters (see Fig. 7.9(a)) are $d = 31$ mm, $g = 2.5$ mm, $W_1 = 2$ mm, and $W_2 = 5$ mm.

The Figures (Figs. 7.20(a) and 7.20(b)) show that for an input power level increasing from −14 dBm to −5 dBm the amplitudes of the transmission and reflection coefficient at resonances decrease a bit. Up to -8 dBm the values stay more or less constant. This is due to the large increase of the real part of the rectifier's impedance at higher power levels, see Table 7.2. The centre of frequency bands R_1 and R_2, do not shift for both transmission and reflection bands, which makes the system more robust for different power levels. The main reason for choosing the commercially available Schottky diode

HSMS 2820 is its low resistive part of the input impedance at low input power levels. To investigate the effect of the direction of arrival (DoA) of the incident wave, the equivalent circuit model for the loaded GSL-FSS unit cell is used to calculate the reflection and the transmission coefficients for different angles of incidence for a TE polarized wave. The input impedance of the rectifier (see Table 7.2) at an input power level of −5 dBm ($R = 39.73\,\Omega$, $C = 1.21\,\text{pF}$) is used in the simulations. Figures 7.21(a) and 7.21(b) show the reflection and the transmission coefficients as a function of frequency for different angles of arrival.

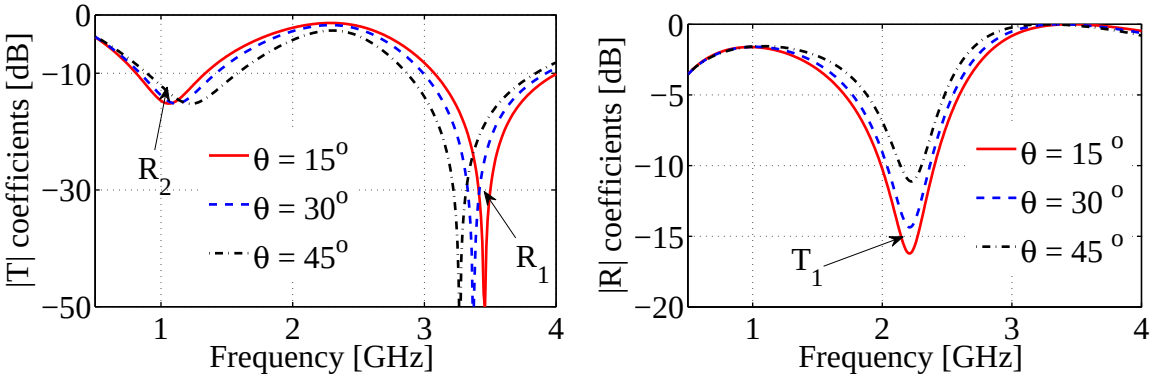


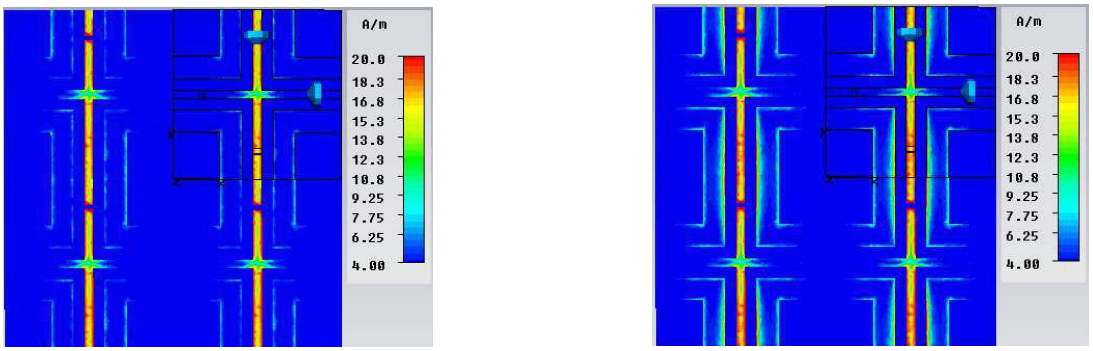
Figure 7.21. Magnitude of the transmission (a) and reflection (b) coefficients versus frequency for different incident angles for $P_{in} = -5$ dBm. The optimized parameters (see Fig. 7.9(a)) are $d = 31$ mm, $g = 2.5$ mm, $W_1 = 2$ mm, and $W_2 = 5$ mm.

Figure 7.21(a) shows that the centre of first reflection band (R_1) shifts to the left (lower frequencies) when the incident angle increases, while the centre of the second reflection band (R_2) shifts to the right. Fig. 7.21(b) shows that the magnitude of the transmission coefficients increases when the angle of incidence increases. Considering -10 dB as a threshold for the reflection and the transmission coefficients amplitude, one can conclude from the figures that for an incident angle less than 45°, the reflection and the transmission resonance frequencies stay within the required threshold. Thus for the Schottky diode (HSMS-2820), and for an input power level increasing from -15 dBm to -5 dBm, the reflection and the transmission frequency bands are robust (i.e. will not be affected) for different angles of incidence.

7.5.1 FSS Current Density

To visualize the resonance phenomena, Figs.7.22(a) and 7.22(b) show the magnitude of the surface current distribution at the centre of the reflection frequency band (1 GHz) (a) and at the centre of transmission frequency band (2.2 GHz) (b), simulated with CST

Microwave Studio. It is clear that for a normal incident, TE polarized wave ($\theta_{inc} = 0°$), the current density is concentrated in the grid and not in the loops, which is desired for power harvesting since the diodes are integrated in this grid. We will validate the possibility to harvest power at the centre of the reflection band ($R_2 = 1$ GHz).



(a) (b)

Figure 7.22. Magnitude of the surface current distribution at the centre of the reflection band $R_2 = 1$ GHz (a) and at the centre of the transmission band $T_1 = 2.2$ GHz (b). The GSL FSS dimensions are: $d = 31$ mm, $g = 2.5$ mm, $W_1 = 2$ mm and $W_2 = 5$ mm. TE with $\theta_{inc} = 0°$

Since, for the designed GSL-FSS, the maximum surface current density is concentrated in the grid, and the simulations indicate that the GSL-FSS is robust for different input power levels and for different angles of incidence (see Figs. 7.20 and 7.21), two gridded square loop frequency selective surfaces have been manufactured and tested. The first prototype is a 3×3 gridded square-loop FSS. The second prototype is a 5×5 gridded square-loop FSS. Both prototypes are shown in Fig.7.23.

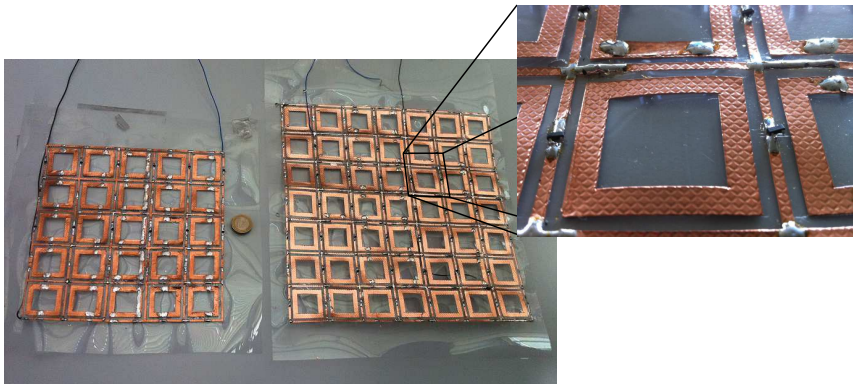


Figure 7.23. Manufactured 3×3 and 5×5 prototypes with $d = 31$ mm, $g = 2.5$ mm, $W_1 = 2$ mm and $W_2 = 5$ mm. Inset showing the soldered diodes in more detail.

As is shown in the figure, both designs are surrounded by a ring of unloaded unit cells to preserve the FSS properties i.e. the coupling between the elements. Both prototypes are printed on a 0.125 mm thick substrate having a relative permittivity equal to 3.0.

7.5.2 RF-Power Transport Measurement Results

The manufactured prototypes are loaded with Schottky diodes, 12 Avago HSMS-2820 [43] diodes for the 3×3 FSS and 30 diodes for the 5×5 FSS. The diodes are placed in the vertical direction to harvest power from TE polarized incident waves. The DC output voltage is measured across an open circuit ($R_L = \infty$) and across a $1\,k\Omega$ resistor ($R_L = 1\,\mathrm{k}\Omega$) connected to the top-left and bottom-right corners with the aid of the DC collection lines as shown in Fig. 7.2(a).

The power density S_{RF} available at a distance R from a transmitter with a transmitted power P_t and antenna transmit gain G_t is:

$$S_{RF} = \frac{P_t G_t}{4\pi R^2} \tag{7.24}$$

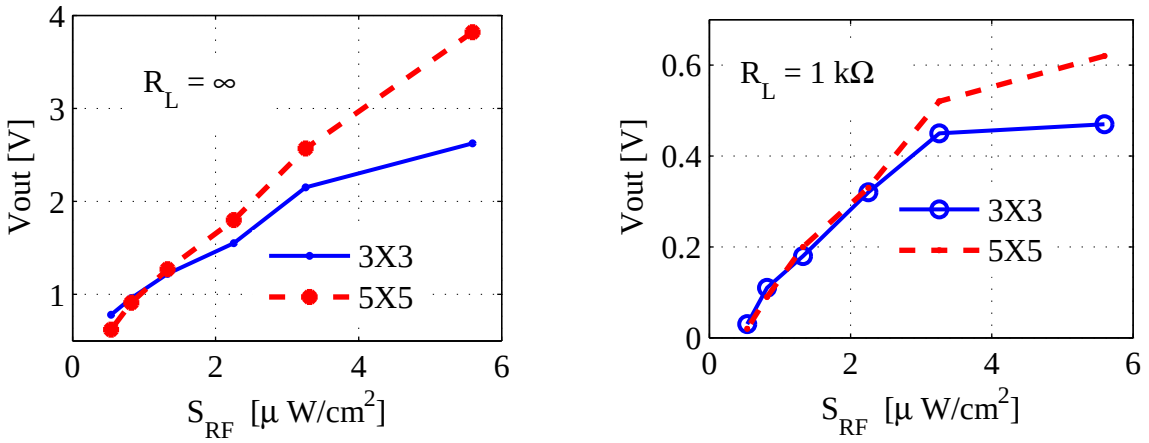
The RF power captured by the FSS surface P_{RF} is assessed using the following equation,

$$P_{RF} = S_{RF} A_{Physiscal}, \tag{7.25}$$

where $A_{Physical}$ is the physical area of the manufactured prototypes. Actually $P_{RF} = S_{RF} A_e$, where A_e is the effective area of the FSS. The physical area will in general be larger than the effective one. By approximating A_e by $A_{Physical}$, we will ensure a conservative assessment of the RF-to-DC power conversion efficiency. The physical area is calculated as follows: $A_{Physical} = ((M+1)p + g + W_1)^2$ (see Fig. 7.9(a)), where M is the number of active cells in the grid. $M = 3$ for the 3×3 FSS and $M = 5$ for the 5×5 FSS.
The RF-to-DC Power Conversion Efficiency (PCE) is then defined as:

$$\eta_{RF/DC} = \frac{V_{out}^2}{R_L} . \frac{1}{P_{RF}}, \tag{7.26}$$

where V_{out} is the measured output voltage across the load resistance R_L. Figure 7.24 shows the open-circuit output voltage (a) and the output voltage across a $1\,\mathrm{k}\Omega$ load (b) as a function of power density available at the FSS surface. It is observed from Fig. 7.24 that both prototypes (3×3 and 5×5) are able to produce an open-circuit output voltage higher than 1V at an available power density larger than $1\mu\,\mathrm{W/cm^2}$.



(a) (b)

Figure 7.24. Measured open circuit output voltage (a), and output voltage across 1 kΩ load (b) as a function of the power density S_{RF}.

Figure 7.25 finally shows the calculated RF-to-DC power conversion efficiency as a function of the input power density. It is shown in the figure that the 3×3 fabricated FSS prototype is more efficient than the 5×5 fabricated FSS prototype. The maximum RF-to-DC conversion efficiency for both prototypes is found at an available power density of 3.2 μ W/cm^2, and is equal to 25 % for the 3×3 prototype and to 16 % for the 5×5 prototype. This is believed to be due to the larger losses and due to the diode cascading saturation effect as previously noticed is Chapter 2 (see Fig. 2.15).

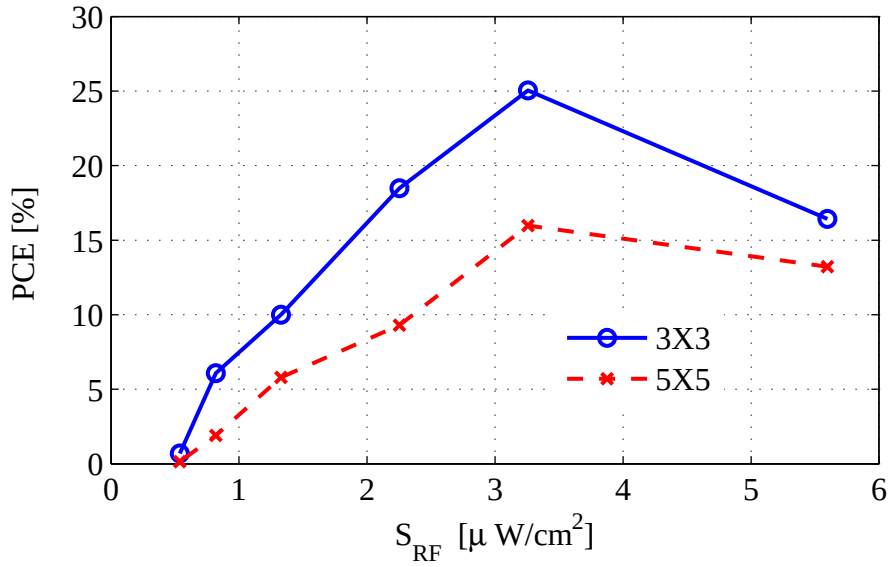


Figure 7.25. RF-to-DC Power Conversion Efficiency (PCE) versus the available power density (S_{RF}).

7.6 Conclusions

The concept, modeling, design, realization and measurement of frequency selective surfaces to harvest RF power is demonstrated in this chapter. The resulting harvester can potentially capture arbitrarily polarized waves, at two different frequency bands, namely the transmission and the reflection bands. A simple and easy-to-implement equivalent circuit model, extended for the insertion of linear lumped components, is investigated, validated and employed for the design. A maximum RF-to-DC power conversion efficiency of 25% at $3.2\mu\,\mathrm{W/cm^2}$ available power density is measured for a 3×3 elements harvester. When the size of the FSS is increased to 5×5 elements, the maximum RF-to-DC power conversion efficiency at the same power density decreases to 16 %, the reason for the decrease in the power conversion efficiency is currently under investigation. This effect is assumed to be due to the the diode cascading saturation effect. The main potential of the novel RF harvester is that it can be implemented for large area RF power harvesting applications. The novel application is suitable for an energy-harvesting glass that can be used in buildings. It could also be implemented in wall papers.

Chapter eight

Conclusions and Recommendations

This thesis addresses the analysis and design of far-field RF power harvesting devices. The potential application of this technology is envisaged in powering wireless sensors that are used in smart buildings. The 'smartness' is found in measuring parameters like e.g. light, temperature and presence of people and acting upon the gathered information (switching off light, turning down heating) thus saving energy consumption. Powering these sensors wirelessly is a challenge due to the restricted transmit power levels allowed. Therefore a highly efficient rectenna design is crucial. It is then important to develop accurate and easy to implement design models for rectenna subsystems, including the rectifier and the antenna. In addition, it is essential to develop a design strategy that copes with the low power density constraint.

A new design strategy has been developed. It is found that we need to optimize a rectenna for a low input power level (-10 dBm), conjugate match the antenna to the rectifier and design the rectifier circuit for a minimum real part and maximum imaginary part $|X_{rec}| \gg |R_{rec}|$. In implementing this design strategy, it has been found that the antenna can be miniaturized up to a level that the radiation resistance still dominates the loss resistance. Consequently, a compact and highly efficient rectenna system that is matched for a broad input power band can be designed.

An analytical model to determine the input impedance and the output voltage of a high-frequency, rectifying Schottky diode is discussed. Based on this model, the influence of the contact resistance and the saturation current on the DC output voltage is analyzed. It has been shown that a higher saturation current and/or a lower contact resistance will result in a higher output voltage. Based on these observations, the commercially available Schottky diode HSMS-2852 has been selected for the rectenna design.

8.1 Conclusions

The first rectenna system presented in this thesis is a 50 Ω rectenna system, where a modified 50 Ω Yagi-Uda antenna is introduced to realize a reasonable gain (4 dBi) over a frequency bandwidth covering the Digital Television broadcasting frequency bands (475 MHz - 794 MHz). The modified Yagi-Uda antenna is implemented in the rectenna design, using a lumped-element matching network to match the antenna impedance to the complex input impedance of a voltage doubling rectifier. The suggested 50 Ω rectenna system shows an increased power conversion efficiency compared to state-of-the-art results reported in the open literature, see Table 8.1.

Table 8.1. Comparison of the 50 Ω rectenna system with results published in the open literature that are based on commercially available discrete diodes.

Ref.	Pin [dBm]	Frequency [MHz]	Resistive Load kΩ	Discrete rectifier	PCE (%)
[66]	-10	866.5	3	HSMS285C	24%
[19]	-10 -20	1800	2.4	HSMS2850	38% 15%
[67]	-10	830	10^4	HSMS286Y	44%
[68]	-20	850 1850	-	SMS7630	14% 13%
This work (50Ω rectenna)	-10 -20	868	10	HSMS2852	49.7% 31.6%

The second rectenna system presented in this thesis deals with antenna structures intended for a direct conjugate matching to the rectifier, thus omitting an impedance matching network, that will lead to a higher power conversion efficiency (less loss) and a more compact design (fewer lumped elements). For this reason, different antenna structures are investigated that are capable of being tuned to the required complex impedance by changing some geometrical features. The validity range (length and width) of the expression for the impedance of a strip dipole antenna has been extended by deriving new fitting equations. Next, the accuracy of an existing strip-folded dipole antenna has been improved by using the improved strip-dipole antenna model. Evolving from this latter antenna, a new, electrically small antenna type has been invented that allows for tuning the complex input impedance and the radiation efficiency independently.

The new modified loop antenna is used for a rectenna design, and is co-designed with a commercially available Schottky diode based rectifier and with an integrated custom-made rectifier. Both realized rectennas demonstrate characteristics better than the state

of the art as reported in the open literature, see Tables 8.2 and 8.3.

Table 8.2 gives a short overview of the state-of-the-art results for rectennas as reported in the literature that use commercially available diodes for RF-to-DC power conversion.

Table 8.2. Overview of rectenna performances with commercially available diodes in the literature.

Ref.	Pin [dBm]	Frequency [MHz]	Load R_L [Ω]	$\frac{Size}{\lambda^2}$	PCE (%)
[19]	-10 -20	1800	2.4×10^3	-	38% 15%
[67]	-10	830	1×10^7	0.027	44 %
[96]	-20	550	2×10^3	0.035	18 %
[20]	0	1960	460	0.192	54 %
This work (Conjugate-complex rectenna)	-10 -20	868	10×10^3	0.028	54.7% 33.8%

Table 8.3 gives an overview of the state-of-the art results for rectennas as reported in the literature that use custom-made integrated rectifiers.

Table 8.3. Overview of rectenna performance with custom-made integrated rectifiers as reported in the literature.

Ref.	This work (Conjugate-complex rectenna)	[101]	[102]	[103]
Technology	90 nm	0.18 μm	250 nm	90 nm
Frequency	868 MHz	970 MHz	906 MHz	915 MHz
Sensitivity	-27 dBm @ 1 V	-17.7 dBm @ 0.8 V	-22.6 dBm @ 2 V	-24 dBm @ 1 V
Peak PCE (%)	36.6 @ -17 dBm	37 @ -18.7 dBm	30 @ -8 dBm	11 @ -18.83 dBm
$\frac{Size}{\lambda^2}$	0.018	0.039	0.027	-
Measured distance	27 meters @ 1.78 w	-	15meters @ 4 w	-

As a proof of principle, a complete rectenna system is presented where a commercially available power management circuit is used to power a temperature and humidity sensor. The average DC output power and the system's RF to DC conversion efficiency are measured, and compared to those measured for a commercially available rectenna

system operated under the same conditions. It is shown that the commercially available rectenna system fails to operate at low power levels (≤ -6dBm) while the presented system can operate at an RF input power level as low as -13 dBm. At an input power level of 0 dBm, the power conversion efficiency of the commercially available system does not exceed 19 %, while the power conversion efficiency of the presented system reaches 40 %, resulting in more than 20 % improved power conversion efficiency.

Finally, for large-area RF power harvesting applications, the concept, modeling, design, realization and measurement of frequency selective surfaces equipped with Schottky diodes to harvest RF power is demonstrated. A simple and easy-to-implement equivalent circuit model, extended for the insertion of linear lumped components, is investigated, validated and employed for the design. A 3×3 and a 5×5 RF harvesting FSS have been designed, fabricated and measured. An RF to DC conversion efficiency of 25 % for the 3×3 RF harvester and 16% for the 5×5 RF harvester are measured at an incident power density of $3.2\,\mu\mathrm{W/cm^2}$. The novel application is suitable for an energy-harvesting glass that can be used in buildings. It could also be implemented in wall papers.

8.2 Recommendations

In this thesis different rectenna topologies and different methods to match the impedance of the rectifier to that of the antenna have been reported. In addition, harvesting RF power for large areas using frequency selective surfaces was introduced. Possible directions for future research include:

- **Power and data transfer**: The introduced rectenna topologies collect the RF power and convert it into DC power. From a system perspective, a second antenna is needed for communication purposes, i.e. transfer of data. A dual port antenna [129] can be used for data and power transfer, one port can be used for communication purposes and the second port can be used for RF power harvesting. Also the rectifier can be used as a matching element for the communication port [130].

- **Environment independent rectennas**: The developed antennas in this thesis are characterized for a free-space environment. However, when the antenna is brought close to a metallic ground plane, its input impedance will deteriorate due to the coupling with the antenna image. Consequently, the matching between the antenna and the rectifier will deteriorate and the power conversion efficiency will decrease. To overcome this negative effect, Metamaterial (MTM) structures can be employed to design environment independent rectennas. Adding an MTM layer between the existing antenna and the metal object may maintain the antenna matching, and enhance the radiation characteristics. The properties of the

electromagnetic band gap (EBG) will make it possible to keep the dimensions of this layer small [131].

- **RF power harvesting using Frequency Selective surfaces**: We have experimentally verified in Chapter 7 that the use of Frequency Selective Surfaces to harvest RF power is feasible. We do need to investigate the relation between size and power conversion efficiency. Further, for the RF power harvesting glass application, optical transparency is needed. An optically transparent conductive film manufactured with a silver grid layer (AgGL) is used in [109] where an antenna with a radiation efficiency better than 85% is designed and measured. The reported optical transparency is 81.3%. The same technology could be used to develop the transparent RF power harvesting glass. Ideally, printed diodes should be available to fully exploit these advantages.

Appendix A

King-Middleton Second Order Method Investigation

A.1 King-Middleton Second-Order Method

The King-Middleton method is an approximate solution procedure for the second order Hallen's integral equation [70]. This method introduces two empirically found fitting expressions to calculate the input impedance of a cylindrical dipole antenna. The input resistance and the input reactance are calculated using Eq. (A.1) and Eq. (A.2), respectively.

$$R(kL, \frac{a}{\lambda}) = \sum_{m=0}^{4} \sum_{n=0}^{4} a_{mn} \left(kL\right)^m \left(\frac{a}{\lambda}\right)^{-n}, \tag{A.1}$$

$$X(kL, \frac{a}{\lambda}) = \sum_{m=0}^{4} \sum_{n=0}^{4} b_{mn} \left(kL\right)^m \left(\frac{a}{\lambda}\right)^{-n}, \tag{A.2}$$

where a_{mn} and b_{mn} are the coefficients shown respectively in Tables A.1 and A.2 corrected as indicated in [132]. These equations are valid for:

- $1.3 \leq kL \leq 1.7$, where k is the wave number and L is the half-length of the dipole.
- $0.0016 \leq \frac{a}{\lambda} \leq 0.01$, where a is the dipole radius and λ is the wavelength.

Table A.1. a_{mn} coefficients to be used in Eq. (A.1) to calculate the input resistance of a cylindrical dipole antenna.

m\n	0	1	2	3	4
0	0.484315e4	-0.475502e2	+0.237406	-0.517831e-03	0.387125e-6
1	-0.137680e5	0.134736e3	-0.671429	0.146303e-2	-0.109502e-5
2	0.147939e5	-0.144272e3	0.717576	-0.156154e-2	0.116801e-5
3	-0.716807e4	0.699523e2	-0.346901	0.753029e-3	-0.562448e-6
4	0.134303e4	-0.130631e2	0.644693e-1	-0.139347e-3	0.103804e-6

Table A.2. b_{mn} coefficients to be used in Eq. (A.2) to calculate the input reactance of a cylindrical dipole antenna.

m\n	0	1	2	3	4
0	-0.644754e4	0.767385e2	-0.360563	0.709234e-3	-0.488904e-6
1	0.189983e5	-0.237542e3	0.110722e1	-0.218661e-2	0.151517e-5
2	-0.209803e5	0.267034e3	-0.124666	0.247753e-2	-0.172674e-5
3	0.102804e5	-0.131818e3	0.619050	-0.123956e-2	0.869298e-6
4	-0.188863e4	0.245077e2	-0.115922	0.233996e-3	-0.161550e-6

A.1.1 Verification of the King-Middleton Equations by the Finite Integrating Technique

To verify the accuracy of the King-Middleton equations, CST Microwave Studio is used to calculate the impedance of a cylindrical dipole of half-length $= L = 0.2\lambda$ and for different cylindrical radii. The results are compared with the results of the King-Middleton expressions. Fig. A.1 shows the input impedance of the cylindrical dipole antenna for a radius of 2 mm $\left(\frac{a}{\lambda} = 0.002,\ \lambda = 1m\right)$.

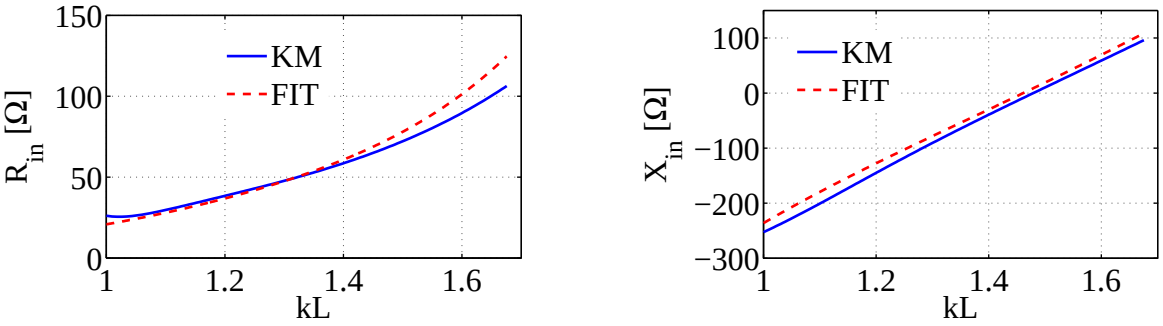


Figure A.1. Real (a) and imaginary (b) parts of the input impedance of a cylindrical dipole antenna calculated by the King-Middleton (KM) second order equations and by the Finite Integration Technique (FIT). Cylindrical radius = 0.002 m, $\left(\frac{a}{\lambda} = 0.002,\ \lambda = 1m\right)$.

It is clearly indicated in the figure that the King-Middleton equations can accurately predict the impedance of a thin cylindrical dipole with less than 10 % difference between the results calculated using the presented expressions and the results calculated using the FIT.

The radius of the cylindrical dipole in increased to 6 mm $\left(\frac{a}{\lambda} = 0.006\right)$ and 10 mm $\left(\frac{a}{\lambda} = 0.01\right)$ and the input impedance is calculated using the King-Middleton equations and the Finite Integration technique. The results are plotted in Fig. A.2 for the radius of 6 mm and in Fig. A.3 for the radius of 10 mm. It is clear from the figures that when the radius of the cylindrical dipole increases, the deviations between the results of the FIT and the results of the KM expressions increase, which validates that the King-Middleton equations are most accurate for thin cylindrical dipoles $\left(\frac{a}{\lambda} < 0.005\right)$.

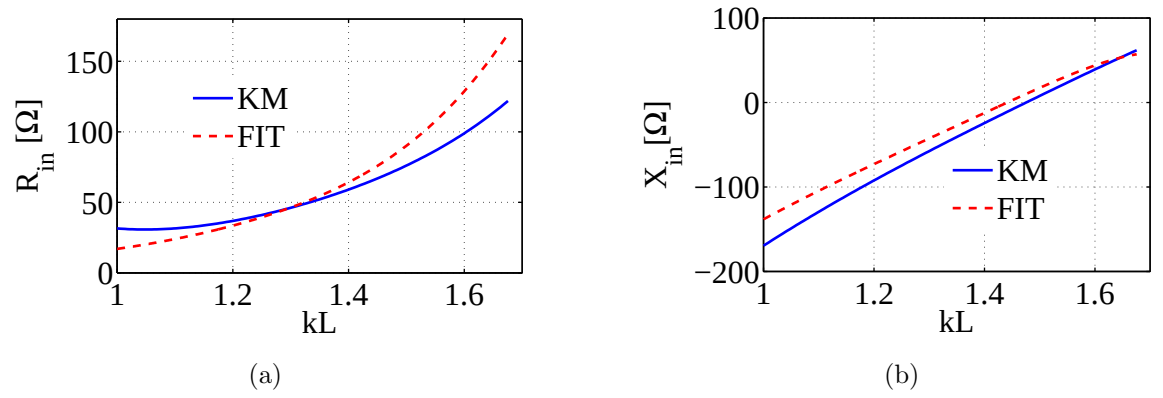


Figure A.2. Real (a) and imaginary (b) parts of the input impedance of a cylindrical dipole antenna calculated by the King-Middleton (KM) second order equations and by the Finite Integration Technique (FIT). Cylindrical radius = 0.006, λ = 1m.

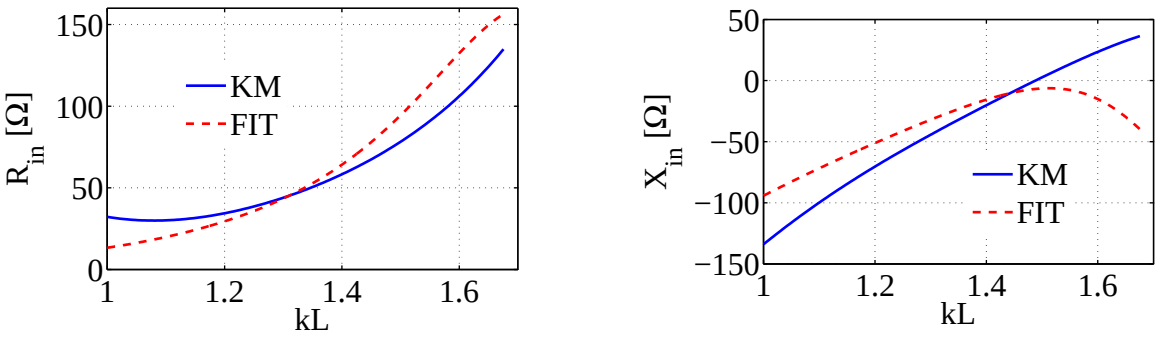


Figure A.3. Real (a) and imaginary (b) parts of the input impedance of a cylindrical dipole antenna calculated by the King-Middleton (KM) second order equations and by the Finite Integration Technique (FIT). Cylindrical radius = 0.01, $\lambda = 1$m.

A.2 Modified King-Middleton Equations

To accurately calculate the impedance of a cylindrical dipole correction factors are introduced into the King-Middleton equations. One for both the real part and one for the imaginary part of the input impedance. The corrections are based on fitting FIT results for an antenna of different radius and length at a frequency of 300 MHz.

A.2.1 Modified King-Middleton Equations - Real Part

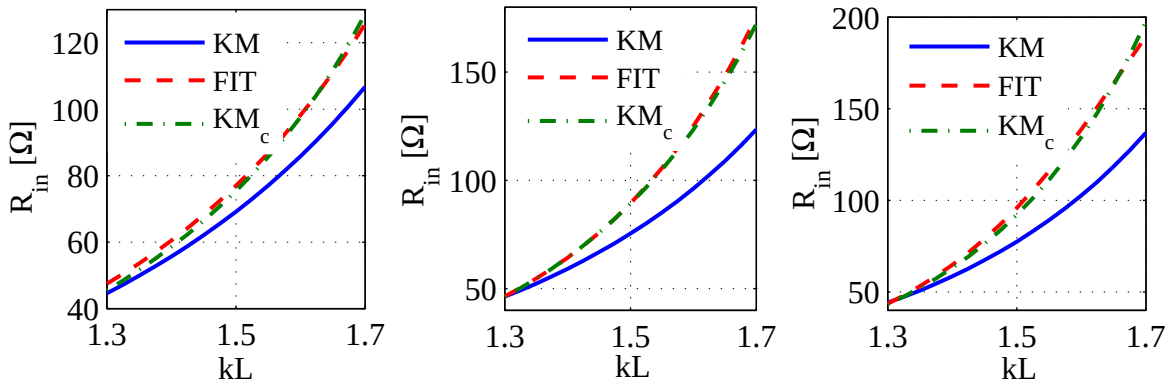
The correction factor of the real part of the input impedance of a cylindrical dipole is expressed by:

$$RC(kL, \frac{a}{\lambda}) = \begin{cases} \alpha e^{\beta\left(\frac{a}{\lambda}\right)} \left(\sum_{m=0}^{4} \sum_{n=0}^{4} a_{mn} (kL)^m \left(\frac{a}{\lambda}\right)^{-n}\right) & \text{for } a < 0.008\lambda \\ \\ \frac{1}{\gamma e^{\delta(kl)}} \left(\sum_{m=0}^{4} \sum_{n=0}^{4} a_{mn} (kL)^m \left(\frac{a}{\lambda}\right)^{-n}\right) & \text{for } a \geq 0.008\lambda \end{cases} \tag{A.3}$$

$$\text{where} \begin{cases} \alpha & = 0.9204 e^{(0.08023(kL))} + 2.565.10^{-7} e^{(7.498(kL))}, \\ \beta & = 35.16 \sin\left(3.393\,(kL) + 8.036\right), \\ \gamma & = 3.657 \sin\left(168.7\left(\frac{a}{\lambda}\right) + 0.02044\right), \\ \delta & = 1.004 \sin\left(189.8\left(\frac{a}{\lambda}\right) + 3.063\right). \end{cases}$$

The coefficients a_{mn} are shown in Table A.1. These modified equations have the same validity ranges as the King-Middleton equations. Fig. A.4 shows the real part of the input impedance as a function of the electrical half-length kL, for a cylindrical dipole

of half-length 0.237 m ($L = 0.237\lambda$) for different dipole radii. The resistive part of the input impedance is calculated using the King-Middleton (KM) equations (solid curve), the corrected King-Middleton (KM_c) equations (dashed-dotted curve) and the Finite Integration Technique (FIT) (dashed curve). It is clearly indicated in the figures that the results of the KM equations deviate from the result of the FIT especially when the electrical length of the cylindrical dipole increases. Moreover, it is shown that when the correction factor is introduced to the the KM equations, the results of the corrected KM (KM_c) equations match the results of the FIT.



(a) $a = 0.0016\ m$ (b) $a = 0.006\ m$ (c) $a = 0.01\ m$

Figure A.4. Real part of the input impedance of a cylindrical dipole of length 0.237 m for different dipole radii calculated using the King-Middleton (KM) Method, the Finite Integration Method (FIT) and the corrected King-Middleton (KM_c).

A.2.2 Modified King-Middleton Equations - Imaginary Part

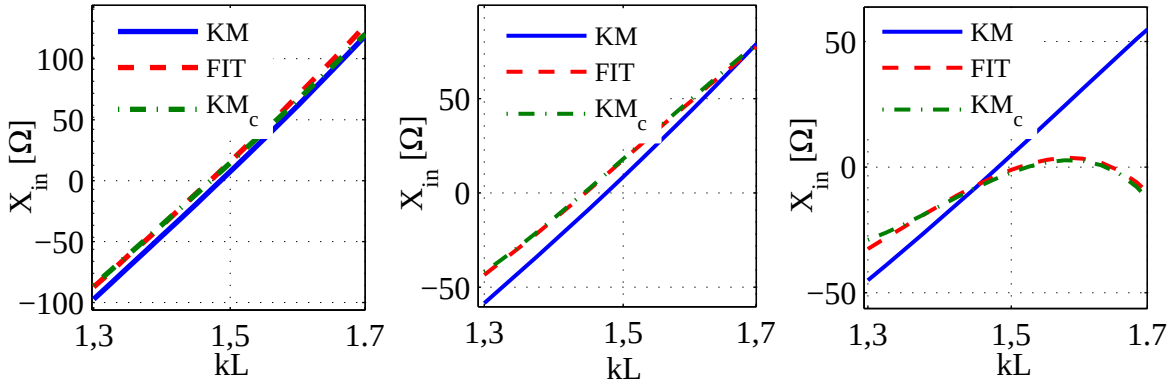
The modified imaginary part of the King-Middleton equation is represented in Eq. (A.4).

$$XC(kL, \frac{a}{\lambda}) = \left(\sum_{m=0}^{4} \sum_{n=0}^{4} b_{mn} (kL)^m (\frac{a}{\lambda})^{-n} \right) + p_0 + p_1 \left(\frac{a}{\lambda} \right) + p_2 \left(\frac{a}{\lambda} \right)^2 \tag{A.4}$$

$$\text{where} \begin{cases} p_0 &= -0.01716 e^{(4.445(kL))} + 9.986 e^{(0.2436(kL))}, \\ p_1 &= 9.107.10^{+9} e^{(-12.47(kL))} + 0.6792 e^{(5.894(kL))}. \\ p_2 &= -27.13 e^{(6.616(kL))}. \end{cases}$$

The coefficients b_{mn} are shown in Table A.2. The imaginary parts of the input impedance of the same cylindrical dipole investigated in the previous section, for different dipole radii as a function of electrical length, are shown in Fig. A.5.
It is shown in the figure that the results of the corrected KM_c expressions overlap the results of the Finite Integration Technique, while the results of the KM expressions

clearly deviate from the results of the FIT especially for large radius cylindrical dipole. To quantify the deviations between the FIT and the KM equation in terms of power wave reflection coefficients, the Figure of Merit introduced in Sec. 1.6 will be used in the following section.

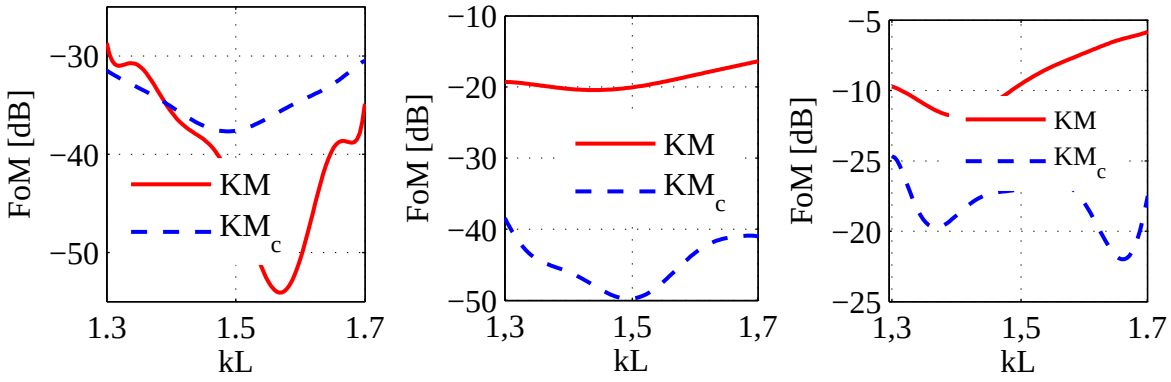


(a) $a = 0.0016\ m$ (b) $a = 0.006\ m$ (c) $a = 0.01\ m$

Figure A.5. Imaginary part of the input impedance of the dipole antenna for different dipole radii using the King-Middleton (KM) Method, Finite Integration Method (FIT) and the corrected King-Middleton (KM_c) .

A.2.3 Comparison Between KM and Corrected KM

From Eq. (1.2), the power wave reflection coefficients due to the deviation between KM equations and FIT is calculated. Using the same equation, the deviations between the corrected KM_c and the FIT are also calculated for the results shown in Fig. A.4 and Fig. A.5. The Figure of Merit (FoM) results are shown in Fig. A.6.



(a) $a = 0.0016\ m$ (b) $a = 0.006\ m$ (c) $a = 0.01\ m$

Figure A.6. Figure of Merit (FoM) to calculate the deviations between KM and FIT and between (KM_c) and FIT as a function of the electrical length for different dipole radii.

It is shown in the figures that the results of the KM_c expressions when compared to the results of the FIT are at least 10 dB more accurate than the KM expressions when compared to the same results (FIT) for larger dipole radii (a = 0.006 m and a = 0.01 m). To validate the modified King-Middleton (KM_c) expressions for different half-lengths, three different cylindrical dipoles are investigated where the input impedance is calculated using the KM expressions, the (KM_c), and the FIT. Fig. A.7 shows the real and imaginary parts of the input impedance of a cylindrical dipole of half-length of 0.119 m and a radius of 2.5 mm. It is shown in the figure that the results of the (KM_c) match the results of the FIT. The same behaviour is shown in Fig. A.8 and in Fig. A.9. In Fig. A.8, the half-length is set to 8 cm, and the cylindrical radius is set to 2.5 mm. The half-length and the radius of the cylindrical dipole are decreased to 5.5 cm and 1 mm and the results of the input impedance are plotted in Fig. A.9. It is shown in the figures that the modified KM (KM_c) expressions can accurately predict the impedance of a cylindrical dipole.

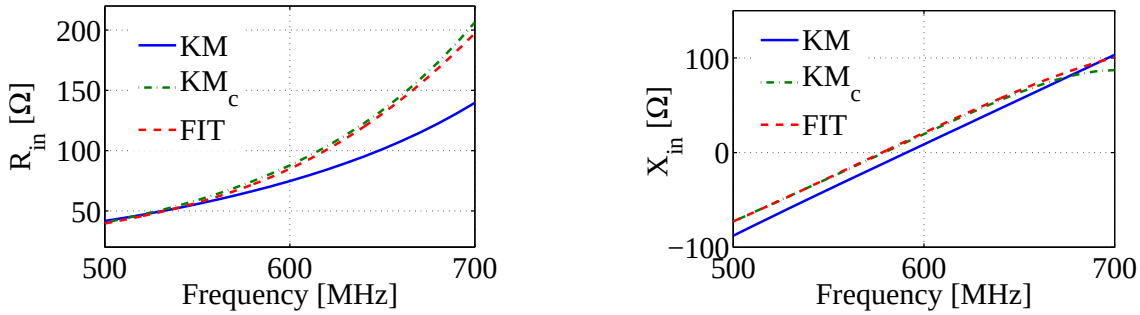


Figure A.7. Real and Imaginary parts of the input impedance, using King-Middleton (KM), corrected King-Middleton (KM_c) and the Finite Integration Technique (FIT). Cylindrical length = 11.9 cm, cylindrical radius = 2.5 mm.

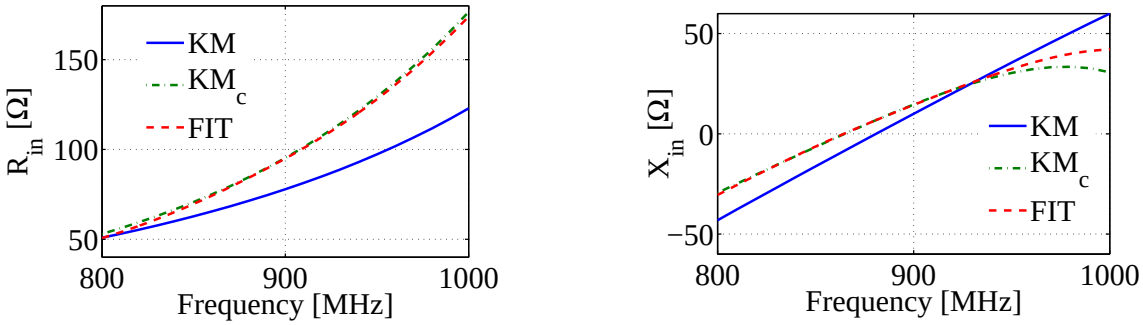


Figure A.8. Real and imaginary parts of the input impedance, using King-Middleton (KM), corrected King-Middleton (KM_c) and the Finite Integration Technique (FIT). Cylindrical length = 8 cm, cylindrical radius = 2.5 mm.

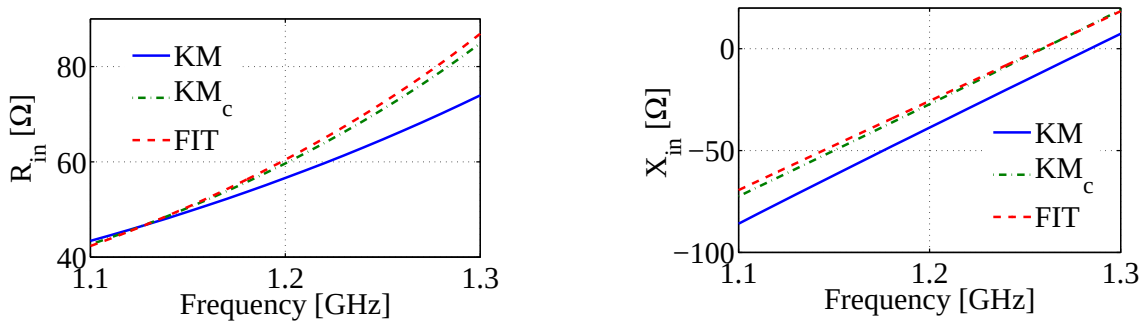


Figure A.9. Real and imaginary parts of the input impedance, using King-Middleton (KM), corrected King-Middleton (KM_c) and the Finite Integration Technique (FIT. Cylindrical length = 5.5 cm, cylindrical radius = 2.5 mm.

A.3 Strip Folded-Dipole Antenna

In this section, the accuracy of the analytical expressions for the strip folded dipole antenna is discussed. Two simulation examples implementing the analytical expressions introduced in Sec. 4.3 at 300 MHz and at 1.5 GHz are presented. Fig. A.10 shows the input impedance (real and imaginary parts) of a strip folded-dipole antenna possessing the following dimensions, $W_1 = 0.04\,\text{m}$, $W_2 = 0.03\,\text{m}$, $d = 0.0125\,\text{m}$, $L = 0.25\,\text{m}$, calculated with and without the correction factor (see Eq. (4.9)) for the inter-element spacing d.
In the same figure FIT-results are shown.

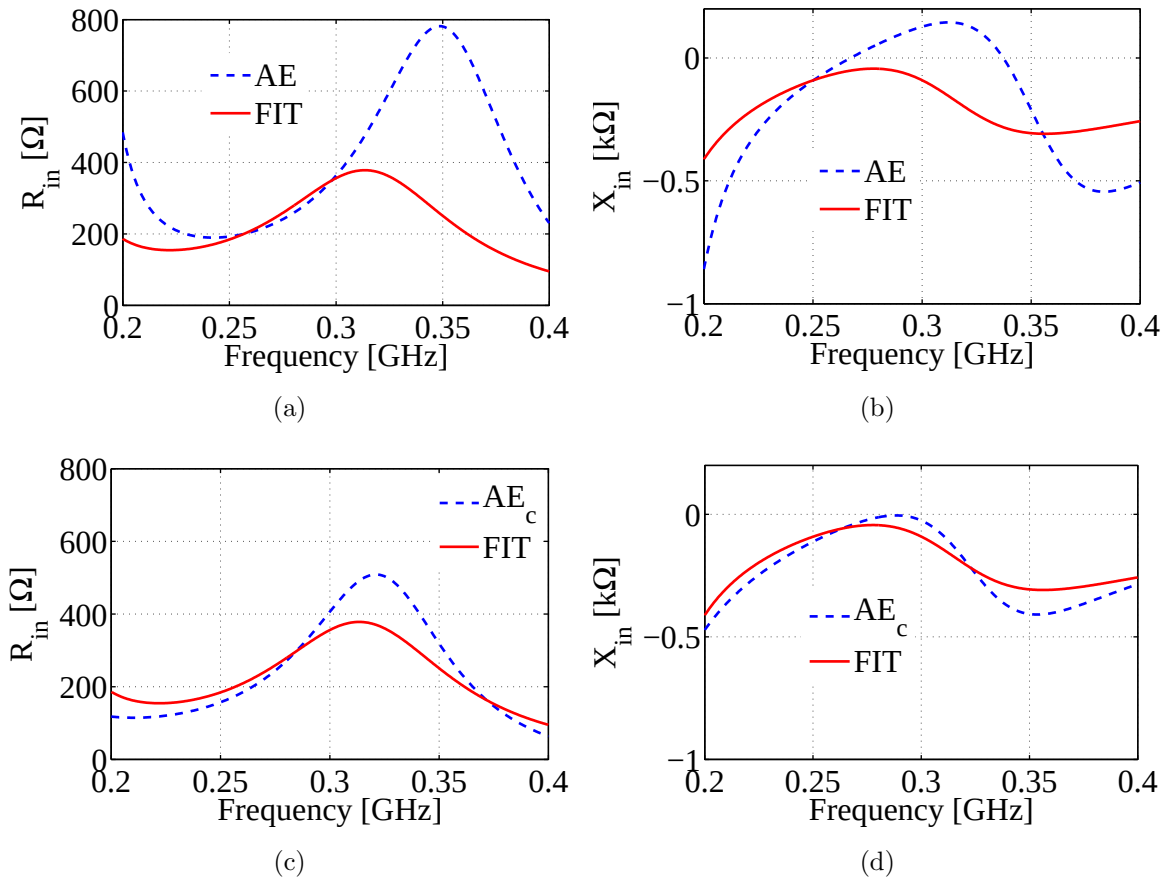


Figure A.10. Real (a) and imaginary (b) parts of the input impedance vs. frequency. And the corrected real (c) and imaginary (d) parts of the input impedance versus frequency. $W_1 = 0.04\,\text{m}$, $W_2 = 0.03\,\text{m}$, $d = 0.0125\,\text{m}$, $L = 0.25\,\text{m}$.

To quantify the improvement of the correction factor in terms of a single Figure of Merit, Eq.(1.2) is used to transform the deviations into power reflection coefficients. Fig. A.11 shows the resulting FoM as a function of frequency with (dashed curve) and without (solid curve) implementing the correction factor in the analytical expressions. Both results are compared to the results of the Finite Integration Technique (FIT). It is clear from the figure that when the correction factor is implemented, the validity range (0.2 GHz - 0.4 GHz) is extended 1.5 times compared to the validity range without implementing the correction factor (0.22 GHz - 0.3 GHz).

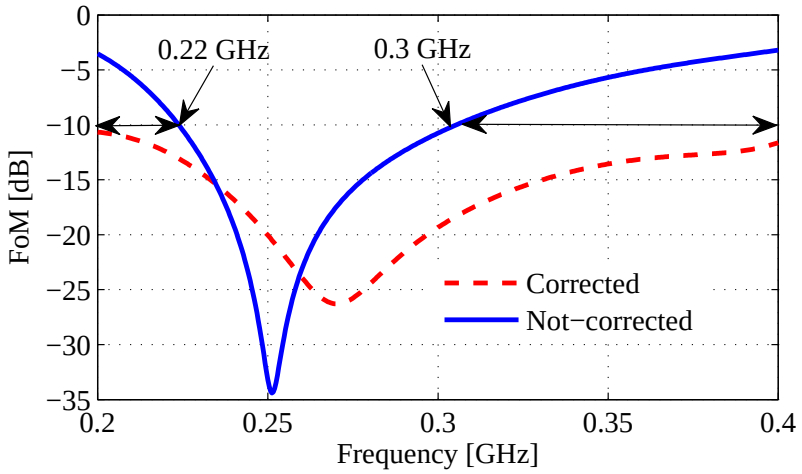


Figure A.11. FoM as a function of frequency due to the mismatch between AE and FIT.

Fig. A.12 shows the input impedance (real and imaginary parts) of a strip folded-dipole antenna possessing the following dimensions, $W_1 = 0.004\,\text{m}$, $W_2 = 0.002\,\text{m}$, $d = 0.004\,\text{m}$, $L = 0.0475\,\text{m}$, calculated with and without the correction factor (see Eq. (4.9)) for the inter-element spacing d. Figs. A.12(a) and A.12(b) show the real and imaginary parts, consecutively, of the input impedance calculated using the analytical expressions presented in Sec. 4.3 without implementing the correction factor for the inter-element distance d. In addition, the Figures show the real and imaginary parts of the input impedance calculated by the FIT using CST Microwave Studio.
The results after using the correction factor are plotted in Figs. A.12(c) and A.12(d). It is indicated in the figures that when the correction factor is implemented, the deviations between the results of the FIT and the results of the analytical expressions decreases. To quantify the improvement when the correction factor is implemented, the FoM is used to transform the deviations into reflection coefficients. Fig. A.13 shows the resulting FoM as a function of frequency with (solid curve) and without (dashed curve) implementing the correction factor in the analytical expressions. It is clear from the figure that when the correction factor is implemented, the validity range (1.22 GHz - 1.88 GHz) is extended more than 50 % compared to the validity range without implementing the correction factor (1.22 GHz - 1.64 GHz).

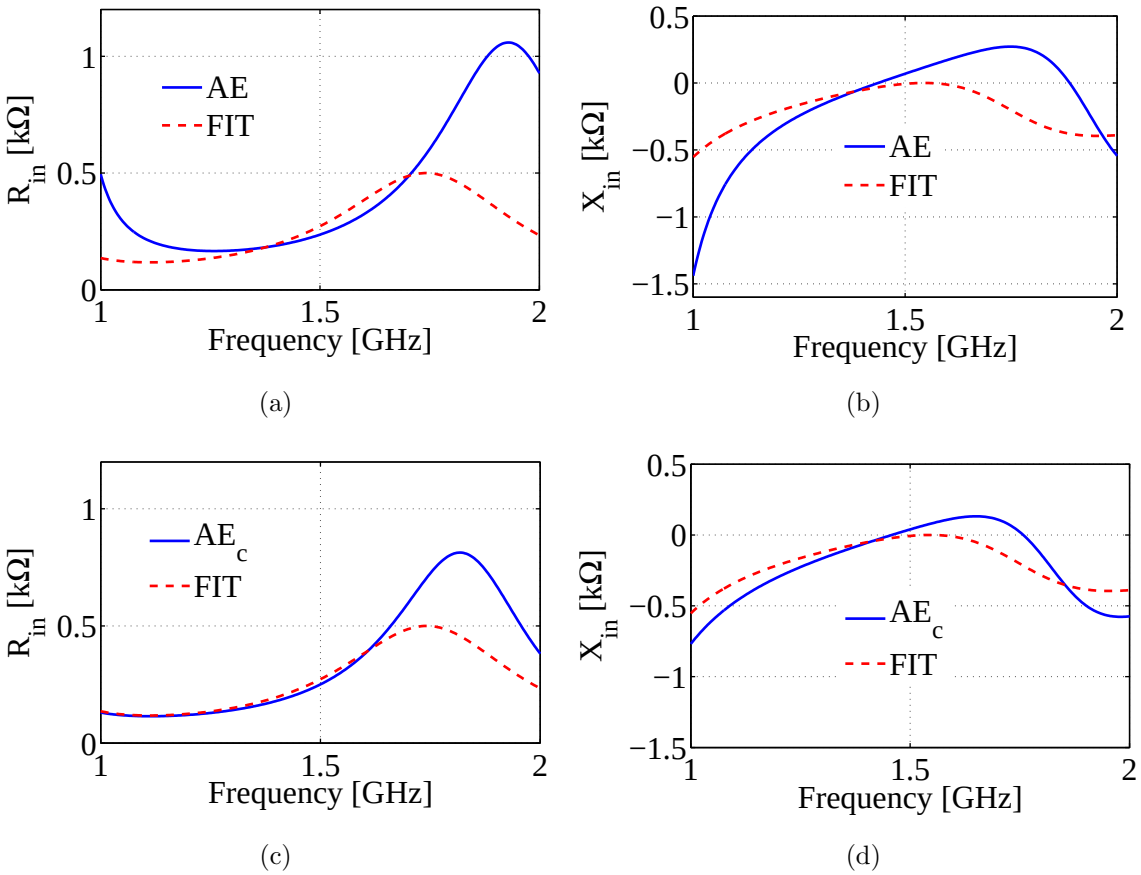


Figure A.12. Real (a) and imaginary (b) parts of the input impedance versus frequency. And the corrected real (c) and imaginary (d) parts of the input impedance versus frequency. $W_1 = 0.004$m, $W_2 = 0.002$m, $d = 0.004$m, L = 0.0475 m.

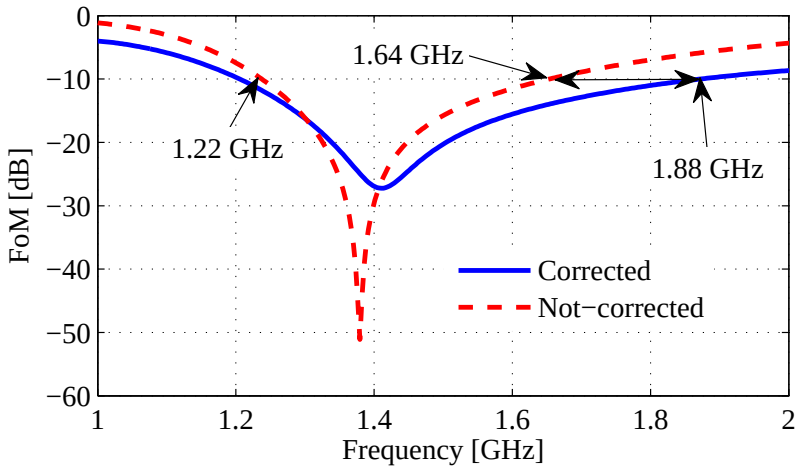


Figure A.13. FoM as a function of frequency due to the mismatch between AE and FIT.

Appendix B

Frequency Selective Surfaces Equivalent Circuit Model

B.1 Correction Factor G

In [119] the first order correction term G, at oblique angle of incidence is obtained as:

$$G(p, d, \lambda, \theta, \varphi) = \frac{A}{B}, \tag{B.1.1}$$

where:

$$A = \frac{1}{2}(1-\beta^2)^2 \left[\left(1 - \frac{\beta^2}{4}\right)(C_{+1} + C_{-1}) + 4\beta^2 C_{+1} C_{-1} \right], \tag{B.1.2}$$

$$B = \left(1 - \frac{\beta^2}{4}\right) + \beta^2 \left(1 + \frac{\beta^2}{2} - \frac{\beta^2}{8}\right)(C_{+1} + C_{-1}) + 2\beta^6 C_{+1} C_{-1}. \tag{B.1.3}$$

In the above:

$$\beta = \sin\left(\frac{\pi w}{2p}\right). \tag{B.1.4}$$

The coefficients $C_{\pm 1}$ are obtained from:

$$C_n = \frac{1}{tS_n} - \frac{1}{|n|}, \qquad n = \pm 1, \pm 2, ... \tag{B.1.5}$$

For a TE polarized incident plane wave:

$$t = j\frac{p\cos(\theta)}{\lambda}, \tag{B.1.6}$$

$$S_n = \frac{\gamma_n}{\gamma_0}. \tag{B.1.7}$$

The propagation constant γ_n is given by:

$$\gamma_n = -j\sqrt{\left(\sin(\theta) + \frac{n\lambda}{p}\right)^2 - 1}, \tag{B.1.8}$$

from which follows that $\gamma_0 = \cos(\theta)$. From B.1.12 and B.1.7:

$$S_n t = -j\sqrt{\left(\frac{p\sin(\theta)}{\lambda} \pm n\right)^2 - \frac{p^2}{\lambda^2}}, \tag{B.1.9}$$

and from B.1.5:

$$C_n{}^{TE} = \frac{1}{\sqrt{\left(\frac{p\sin(\theta)}{\lambda} + n\right)^2 - \frac{p^2}{\lambda^2}}} - \frac{1}{|n|}. \tag{B.1.10}$$

Therefore:

$$C_{\pm 1}{}^{TE} = \frac{1}{\sqrt{\left(\frac{p\sin(\theta)}{\lambda} \pm 1\right)^2 - \frac{p^2}{\lambda^2}}} - 1. \tag{B.1.11}$$

For the dual case of a TM polarized incident plane wave:

$$t = j\frac{p\sec(\varphi)}{\lambda}, \tag{B.1.12}$$

$$S_n t = \gamma_n \gamma_0, \tag{B.1.13}$$

where:

$$\gamma_n = -j\sqrt{\left(\sin(\varphi) + \frac{n\lambda}{p}\right)^2 - 1}, \tag{B.1.14}$$

from which $\gamma_0 = \cos(\varphi)$ is obtained. Therefore:

$$S_n t = \sqrt{\left(\frac{p\sin(\varphi)}{\lambda}\right)^2 + n^2 - \frac{p^2}{\lambda^2}}, \tag{B.1.15}$$

and from B.1.5:

$$C_n{}^{TM} = \frac{1}{\sqrt{\left(\frac{p\sin(\varphi)}{\lambda}\right)^2 + n^2 - \frac{p^2}{\lambda^2}}} - \frac{1}{|n|}, \tag{B.1.16}$$

from which follows that

$$C_{\pm 1}{}^{TM} = \frac{1}{\sqrt{1 - \left(\frac{p\cos(\varphi)}{\lambda}\right)^2}} - 1. \tag{B.1.17}$$

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Curriculum Vitae

Shady Keyrouz was born on January 21st, 1984, in Bsharri, Lebanon. He enrolled in electrical engineering at the Notre Dame University (NDU) in Beirut, Lebanon, in 2003 and obtained his BE degree in 2008. He received his M.Sc. degree in Communication Technology from Ulm University, Germany, in 2010. In the same year he started as a Ph.D.-student at the Electromagnetics group (EM) of the Faculty of Electrical Engineering of Eindhoven University of Technology (TU/e), Eindhoven, The Netherlands. In 2010 he joined Holst Centre/imec as a Ph.D. researcher. His research interests include antenna modeling, rectenna design, wireless power transmission and reflect-array antennas.

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Rectenna materials improving low scale RF power harvesting for embedded miniature sensors

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Abstract

The paper reviews suitable innovative materials used to improve harvesting power through ambient Radio Frequency (RF) needed for standalone embedded sensors. The Rectenna (a combination of a rectifier and antenna) material is one of the main three key elements constituting the scavenging system including the resonator based on the material, the number of multiplier stages and the low pass filter. The system uses electromagnetic waves travelling in ambient air captured through an antenna to an electronic unit and will excite a subsystem to harvest energy. This subsystem is the heart of the concept of power scavenging and is material based concept in most of the cases treated in this paper.

Keywords: Power harvesting, RF, resonator, rectenna, batteryless

1. Introduction

Permanent energy supply is highly needed for battery-less standalone electronic devices. The daily and continuous use of these devices exhibit a real issue with batteries and their fast recharging. Methods for extracting or harvesting energy from the surrounding environment are of great importance because it reduces the maintenance scheduling, transportation, weight and cost of energy needed by the standalone devices. The energy sources of interest are the largely untapped sources of ambient energy resulting from human activity and environmental energy flows in the form of waste heat, movement and vibration.

It is important to remind that power transmission dates back in early work of Hertz who has shown the propagation of electromagnetic waves in air. Tesla followed by transmitting power from one side to another wireless in 1899.

A couple of innovative techniques have already shown a possibility of improvement of scavenged power through RF waves in ambient air. The known three parameters governing the power scavenging are the resonator, the number of multiplier stages and the low pass filter. The resonator or the rectenna shows better performance if specific promising micro-scale energy harvesting standard materials are used e.g. ceramics, single crystals, polymers and composites, and also new engineered materials currently being developed. Resonator can provide increased voltage level with single stage voltage multiplier. Further improvement with this scheme can be obtained by multiple stages. It is also demonstrated that the use of LPF provides the output voltage with little amount of harmonics. Since RF to DC conversion efficiency has emerged as challenging issue, so band pass filter can be used to complete this objective by removing the second and third harmonic at antenna [1].

Our current interest is material. The energy sources of interest to this paper is in the Low power requirements (μW to mW) – for example heat and movement lost from portable electronics and mobile communications. The aim is not to generate large-scale power, but to capture small amounts of ‘energy’ that is ‘wasted’ during industrial and everyday processes. There is currently a great commercial potential for energy harvesting technologies, hence many projects are supported by European companies, from

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automotive to electronics manufacturers as well as European framework 7 programmer [2], [3]. It important to note the amount of harvested power has very low density power ranging from 0.1 μW/cm3 to 0.001 mW/cm3. Fig. 1 shows measured GSM-900 power density levels.

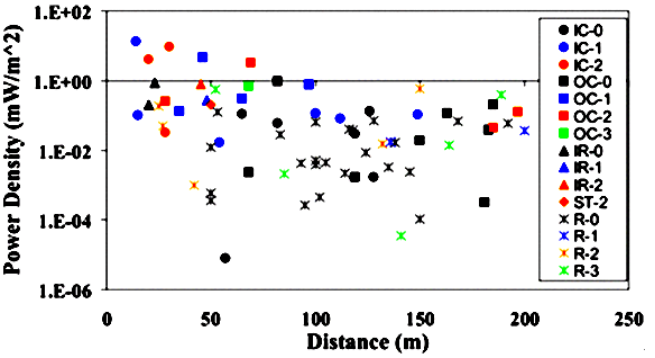
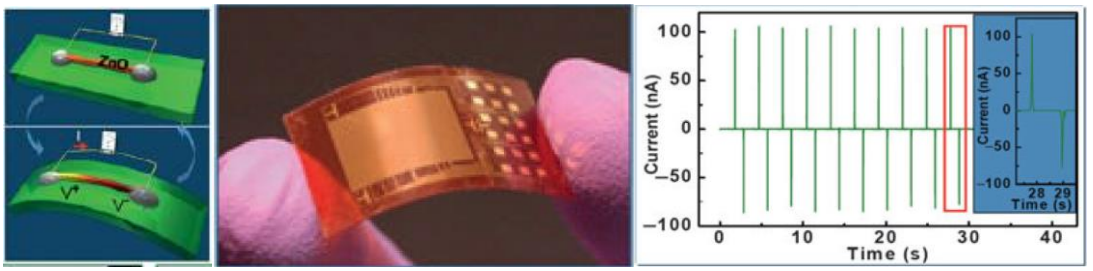


Fig. 1. Measured GSM-900 peak power density levels as a function of distance to the nearest base station. IC=Inner City, OC=Outer Country, IR=IndustRial area, ST=Small Town, R=Rural or country-side are. a: 0=outdoor on the ground, I=outdoors on roof, terrace or balcony, 2=indoors, close to windows, 1.5m or less, 3=indoors, not close to windows [4].

2. Types Materials Used in RF-Based Energy Harvesting

Several promising micro-scale energy harvesting standard materials are used e.g. ceramics, single crystals, polymers and composites, and also new engineered materials currently being developed. Micro- and nano-scale Metal Insulator-Metal (MIM) tunnel diodes are being developed [5] to provide half-wave rectification as part of a "rectenna" energy harvesting system, which includes a radiation-collecting antenna, a rectifying MIM tunnel diode, and a storage capacitor. High-frequency MIM tunnel diodes for power rectification were designed, fabricated and characterized. Planar $Pt/TiO_2/Ti$ stacks are being fabricated to create a diode with highly asymmetric I-V characteristics that has a very low threshold voltage. A MIM tunnel diode is composed of two conductive metal layers separated by a very thin dielectric, and operates on the principles of thermionic emission and quantum tunneling Preliminary results exhibit asymmetric I-V characteristics with threshold voltages of less than 700 mV. Initially this MIM was a poor resistive element because of fabrication issues.

Single and multi-walled carbon nanotube-based technology for radio frequency [6] switch working in the range of 40–60 GHz and fulfilling the specifications were exploited. The first processed component had an operating voltage of 14 V for an ohmic contact in a Nano-Electro Mechanical System (NEMS) tweezer design. High Frequency Simulation Software (HFSS) RF-simulation on an innovative NEMS geometry shows encouraging results with transmission ratios between "on" state and "off" state up to 34% for ohmic-contact switch and 25% for a capacitive-contact switch.



a) b) c)

Fig. 2. a) principle of ZnO with two metal electrode in equilibrium and stretched, b) high output nanogenerator using integration of millions of nanowires transferred to polymer substrate. c) output of 2V and 100 nA are obtained.

Piezoelectric cantilever [7], [8] is used to induce strain via RF signal propagation. Hence, when a piezoelectric cantilever is subjected to RF propagation field, an alternating current is induced in the piezoelectric cantilever causing mechanical movement. Hence, maximum effect is achieved when RF frequency matches the resonant frequency of the cantilever. When a voltage signal of 10 Vpp propagates in the line it sets an alternating current in the actuator electrodes. The effect of piezoelectric material was also exploited in ZnO nanowires laterally bonded on a flexible substrate (Fig. 2). One of the unique advantages of ZnO is that all of the nanowires grow along the *c* -axis (i.e., in the polar direction) and are uni-axially aligned. Once the wires are strained, a macroscopic piezo-potential is created, which can drive a flow of electrons, thus converting mechanical energy into electricity.

On another application [9], RF waves are received from a mobile phone and rectified to generate a DC voltage by a rectenna (a rectifying circuit with an antenna). The layered rectenna is photo-etched from copper-clad FR4 material [10], which allowed for a fast realization of the product. To model the dc output voltage, the incident power received by the patch antenna is measured for a fixed transmit-receive setup. This data is used in the model to predict the dc output voltage of the rectenna. The dc output voltage is predicted and measured in Fig. 3. The system was fabricated with 0.6 _m-CMOS technology, and can supply well-regulated 4.0 V/1mA DC power to the load for 10 ms periodically.

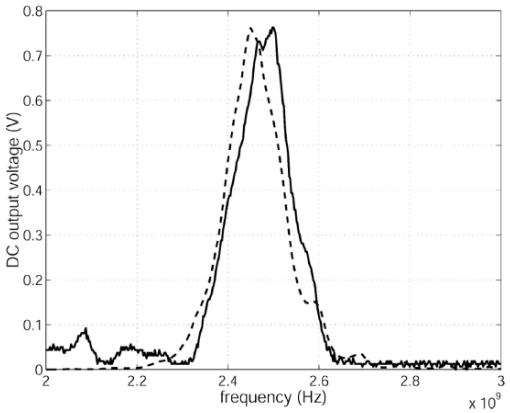
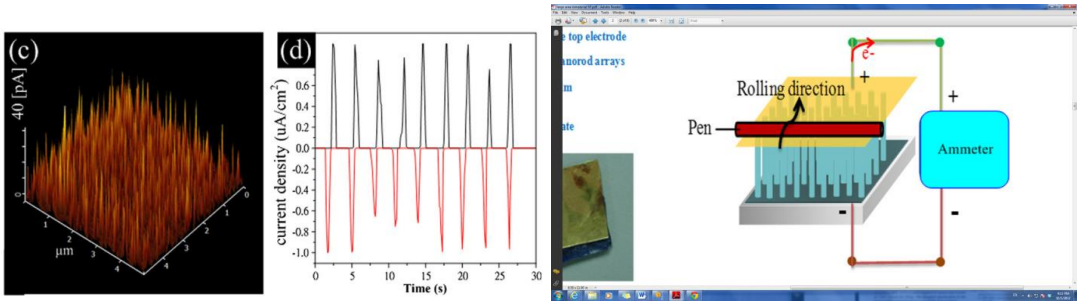


Fig. 3. DC output voltage of the stacked rectenna. Measurement (solid) and model (dash).



a) b) c)

Fig. 4. a) Nanorods grown on, b) current received from 2x2 cm^2 based on concept in c).

ZnO is also used to generated power as follows [11]. An optimal Radio Frequency (RF) sputtering process (sputtering power, sputtering time, and target type) to grow a ZnO seed layer on an Indium Tin Oxide (ITO) glass followed by the hydrothermal growth of a well-aligned perpendicular ZnO nanorod array for the fabrication of a piezoelectric nanogenerator Fig. 3. The change in nanorod morphology, crystallinity, orientation, growth rate, diameter, length, and number density of ZnO nanorods were controlled by modifying various processing parameters. A prior seeding of ZnO nanoparticle with good wurtzite structure and with preferred orientation along the (002) direction of the crystal is critical for the

subsequent growth of well-aligned vertically oriented ZnO nanorods on ITO substrates. A large area (2_2cm^2) of the piezoelectric nanogenerator bimaterial sheet was fabricated by combining the vertically grown ZnO nanorod array electrode with an Au-coated polyurethane (PU) film as the other electrode. By rubbing the two electrodes together, the nanodevice generates an optimal output current density of about $1\mu A/cm^2$ in Fig. 4c).

3. Conclusion

A couple of innovative techniques have already shown a possibility of improvement of scavenged power. The known three parameters governing the power are the resonator, the number of multiplier stages and the low pass filter. The resonator or the rectenna has shown better performance if specific promising micro-scale energy harvesting standard materials are used e.g. ceramics, single crystals, polymers, composites, and new-engineered materials currently under development.
Piezoelectric nanogenerators have been developed that can serve as self-sufficient power sources formicro-/nanosystems. For wurtzite structures that have non-central symmetry, such as ZnO, GaN, and InN, a piezoelectric potential (piezopotential) is created by applying a strain.

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A Novel Ultra-Lightweight Multiband Rectenna on Paper for RF Energy Harvesting in the Next Generation LTE Bands

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A Novel Ultra-Lightweight Multi-Band Rectenna on Paper for RF Energy Harvesting in the Next Generation LTE Bands

Valentina Palazzi, *Student Member, IEEE*, Jimmy Hester, *Student Member, IEEE,* Jo Bito, *Student Member, IEEE,* Federico Alimenti, *Senior Member, IEEE,* Christos Kalialakis, *Senior Member, IEEE*, Ana Collado, *Senior Member, IEEE*, Paolo Mezzanotte, *Senior Member, IEEE*, Apostolos Georgiadis, *Senior Member, IEEE*, Luca Roselli, *Senior Member, IEEE*, and Manos M. Tentzeris, *Fellow, IEEE*

***Abstract*—This paper introduces a novel compact ultra-lightweight multiband RF energy harvester fabricated on a paper substrate. The proposed rectenna is designed to operate in all recently released LTE bands (range $0.79-0.96$ GHz; $1.71-2.17$ GHz; $2.5-2.69$ GHz). High compactness and ease of integration between antenna and rectifier are achieved by using a topology of nested annular slots. The proposed rectifier features an RF-to-dc conversion efficiency in the range of $5-16\,\%$ for an available input power of -20 dBm in all bands of interest, which increases up to $11-30\,\%$ at -15 dBm. The rectenna has been finally tested both in laboratory and in realistic scenarios featuring a superior performance to other state-of-the-art RF harvesters on flexible substrates.**

***Index Terms*—Broadband antennas, conformal antennas, copper adhesive laminate, energy harvesting, flexible electronics, Internet of Everything, Internet of Things, multiband rectifier, rectennas, slot antennas**

I. Introduction

In THE near future, an explosion of the number of devices connected to the Internet is foreseen, which will impose tremendous challenges both in terms of spectrum and power supply [1]- [4]. On the one hand, new communication protocols and wider bandwidths will be released to manage this huge number of heterogeneous users (both humans and machines), with different levels of intelligence and needs [5]. On the other hand, according to the emerging Internet of Things (IoT) paradigm, electronic apparatuses will be deployed almost everywhere. The next generation of sensors and circuits will be literally embedded in the items, so their maintenance will be hindered by the fact that they will be difficult to reach or even to localize. This means that the battery replacement will be difficult or even impractical, so in most of the cases batteryless devices will be the only option. Additionally, these devices have to be conformal to the hosting objects both in terms of shape and cost, while the environmental impact of these distributed and ubiquitous electronics should be minimized in order not to increase the pollution risks and cause a worsening of the quality of human life.

For these reasons, it is important to avoid the traditional battery and to find new efficient and reliable ways to power the circuits from ambient RF sources, both exploiting already available resources through wireless energy harvesting and enabling the intentional wireless power transferring, so as to make them robust and autonomous.

To this end, research is focusing on both the development of low-power sensors and energy harvesting to produce autonomous wirelessly interconnected smart nodes [6]. State-of-the-art dc/dc converters are able to operate for input voltages as low as 70 mV [7], thus thoroughly reducing the amount of energy required for electronics. Thanks to this trend, ambient RF energy harvesting is becoming ever more appealing among the different forms of energy harvesting (e.g., solar, mechanical, and thermal). So far, the scarcity of the available power density has been its main drawback, but both the development of lower power and lower voltage circuits and the increasing number of wireless applications, which in the future will lead to a dramatic increase of the ambient RF energy, are concurrently bridging the gap.

In this scenario, broadband rectennas are the most suitable solutions to realize reliable power supply [8], since they are able to harness a larger part of the available spectrum; plus, the harvester should be easily mounted on commonly used objects, so it is necessary that they are lightweight and conformal.

Recently, numerous high-performing broadband and multi-band rectennas have been proposed (see [9]- [15]), which are able to take advantage of the simultaneous presence of RF power at different frequencies. An in-depth analysis of the rectifier architectures and operating principles was conducted as well, aiming at improving specific properties of the harvesting circuits, such as their bandwidth [16], sensitivity [14] and operating power range [17]. However, further improvements are needed in terms of compactness and conformability to achieve truly practical and scalable solutions.

This paper is a substantial expansion [18] and [19]. In the

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previous works, a single band rectenna was designed and demonstrated. Here, both the antenna and the rectifier have been enhanced in order to allow for multiband operation. Furthermore, a compact printed module integrating the proposed new antenna is demonstrated. In the end, a novel ultracompact multiband rectenna is presented, able to harvest power from all released LTE bands.

This paper is organized as follows. Section II presents the system-level design of the multiband rectenna. Section III describes the low-cost fabrication technology utilized for the rectenna prototype. Sections IV and V describe the design procedure of the proposed harvester for the antenna and the rectifier, respectively. Results discussing the performance of the rectenna in various representative harvesting scenarios are given in Section VI. The conclusions are drawn in Section VII.

TABLE I
RELEASED LTE BANDS (EUROPE)

Band	Uplink (MHz)	Downlink (MHz)	Alias
1	1920-1980	2110-2170	UMTS, IMT2100
3	1710-1785	1805-1880	DCS1800
7	2500-2570	2620-2690	IMT-E2600
8	880-915	925-960	GSM900, EGSM900
20	832-862	791-821	EU Digital Dividend 800 MHz
33	1900-1920	1900-1920	pre IMT
34	2010-2025	2010-2025	IMT
38	2570-2620	2570-2620	IMT-E2600 (duplex spacing)

II. System Level Design

The proposed rectenna is aimed at harvesting energy from all new bands released by the Third Generation Partnership Project (3GPP) [20] , [21], both in uplink and in downlink. A list of the bands adopted in Europe is reported in Table I. It has been noticed that all these bands can be divided into two groups well separated in frequency: the bands 8 and 20 on the one hand, which occupy the range from 790 to 960 MHz, and all others on the other hand, which extend from 1710 to 2690 MHz.

Therefore, an approach based on two antennas has been developed instead of a single broadband rectenna. The rectenna block diagram is reported in Fig. 1. The idea is to use a relatively narrowband antenna tuned around 900 MHz and a broadband one which is able to gather energy from frequencies between 1.71 and 2.69 GHz. Both antennas are connected to two separate rectifiers and then their outputs are summed with a dc combining technique. Finally, an RC low-pass filter isolates the dc component, which can be utilized by practical loads (a low-power sensor, a dc/dc converter, and so on).

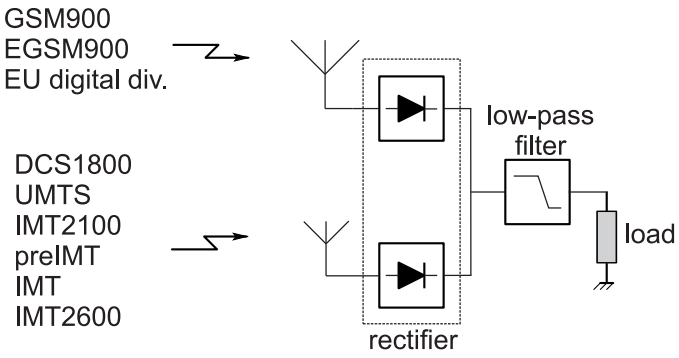


Fig. 1. Architecture of the proposed multiband rectenna.

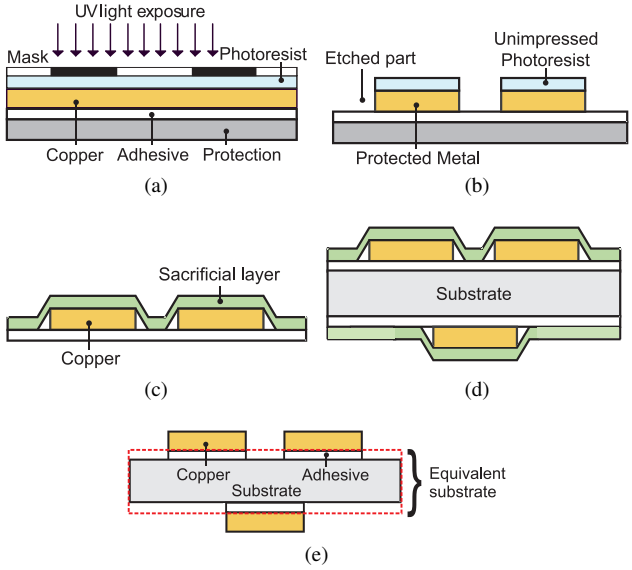


Fig. 2. Adhesive copper laminate technology: fabrication steps. (a) Deposition and selective exposure of photoresist to UV light through a mask. (b) Wet etching of copper. (c) Introduction of a sacrificial layer on the copper side and removal of the protection of the adhesive layer. (d) Transferring of the etched traces to both sides of the substrate. (e) Removal of the sacrificial layer.

Thanks to this approach, the bandwidth requirements for both the antenna and the rectifier are less demanding, thus allowing for a reduction in the complexity of each circuit block. On the other hand, the number of blocks is increased. So, in order not to end up with an increased rectenna area, it is essential to choose the most suitable technology and guarantee a compact arrangement of the utilized components.

III. Adhesive Copper Laminate Technology

The whole circuit has been manufactured on a paper substrate using the adhesive copper laminate technology described in detail in [22]. The main steps of this technology are reported in Fig. 2. It relies on a copper adhesive tape shaped by a photolithographic process [see Fig. 2(a) and (b)] and then transferred to the paper substrate by means of a sacrificial layer [Fig. 2(c)]. Since in the present work a double layer structure is needed, the two layers are processed separately by using two different pieces of copper tape that are then attached to the two sides of the substrate [Fig. 2(d) and (e)] and manually aligned by means of alignment marks. The latter process is estimated to

feature a misalignment tolerance of about 1 mm, which can be considered acceptable in the low-GHz range. In the case of a simple microstrip structure, a slab of copper tape is attached to the bottom side of the substrate so as to obtain a ground plane.

The advantages of the adopted technique with respect to other emerging ones, such as ink-jet printing [23], are higher metal conductivity ($\sigma = 5.8 \times 10^7$ S/m), the possibility to solder surface mount components by means of conventional eutectic alloy while eliminating curing steps, and compatibility with Printed Circuit Board fabrication equipment. The price to pay for this is the presence of a photolithographic step that makes this technique not fully additive.

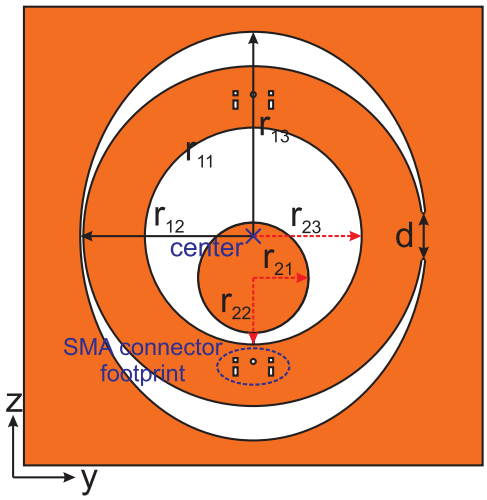
In this context the paper substrate has been chosen for its unique properties of being lightweight, flexible, biodegradable, and intrinsically inexpensive [24], as well as for being an example of unconventional material in full compliance with the IoT paradigm that requires electronic technologies to be as compliant as possible with materials in common objects. The electrical parameters of the adopted substrate, which is a photo-paper from Mitsubishi Electric (model CK-D715), are quoted in [25]. The equivalent composite substrate, including both the paper substrate and the glue layers interposed between each layer of copper and paper, features a permittivity of 2.55, a loss tangent equal to 0.05, and a thickness of 0.37 mm. Despite the fact that the high losses of the substrate represent a substantial limitation to the circuit performance, especially for the rectifier (see Section V), this paper introduces an ultra low-cost solution and can be considered as a proof-of-concept demonstration of the feasibility of autonomous wireless sensor nodes on almost every kind of material.

IV. Antenna Design

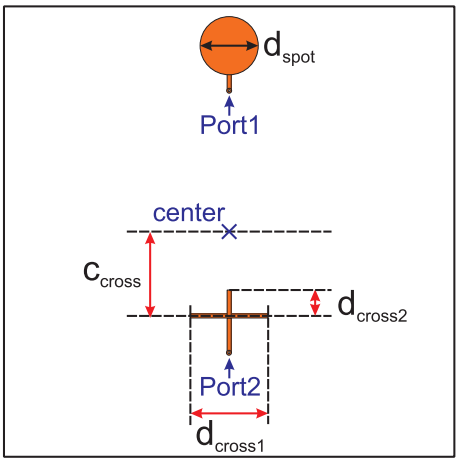
The proposed antenna consists of two annular slots nested one inside the other, according to a technique already demonstrated in [26]. The antenna layout is shown in Fig. 3 and its design was carried out by means of Computer Simulation Technology Studio suite. The antenna input ports are marked with "Port1" and "Port2".

Since the antennas are substantially etched out from the ground plane, they do not add weight to the structure and are compatible with substrates as thin as a sheet of paper. Moreover, slot antennas feature a relatively wide impedance bandwidth due to their low quality factor (see [27]) and, thanks to the exponential decrease of the current density at the borders of the slots, the metal, which surrounds the slots, can be exploited as the ground plane for the circuitry placed on the other side, thus allowing an easy integration between the antenna and the rectifier.

The outer antenna is the tapered annular slot used and analyzed in [18]. The slot, tuned to the band $790 - 960$ MHz, is the result of the intersection of an ellipse and a circle and features a maximum and minimum width of 8.25 and 0.75 mm, respectively. The antenna is fed by means of a proximity coupled microstrip line terminating on a circular stub introduced to match the antenna to a $50 - \Omega$ iimpedance, as shown in Fig. 3(b) (Port1). Since the annular slot features



(a)

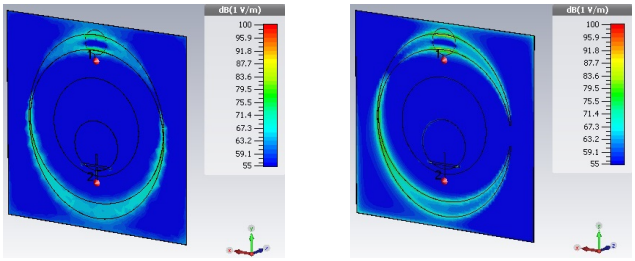


(b)

Fig. 3. Layout of the antenna prototype. (a) Top side. (b) Bottom side. Active area: 11×11 cm^2. The main parameters of the antenna topology are: $r_{11} = 40.75$ mm, $r_{12} = 41.5$ mm, $r_{13} = 49$ mm, $d = 11$ mm, $d_{spot} = 14.2$ mm for the first antenna, $r_{21} = 13.24$ mm, $r_{22} = 15.99$ mm, $r_{23} = 26$ mm, $d_{cross1} = 19$ mm, $d_{cross2} = 6$ mm, $c_{cross} = 21$ mm for the second antenna.

a length equal to λ_g, i.e., the guided wavelength in slot technology at the central frequency, the magnitude of the electric field across the slot features a maximum at the feeding point and at its symmetric point (where the slot width is maximum as well), whereas it is ideally null at the antisymmetric points (approximately where the slot width is minimum). As a consequence, the latter two points can be viewed as virtual grounds and it is possible to replace one of them by means of a real short (marked with "d" in Fig. 3) at the expense of an increased cross-polar component [28]. Fig. 4 reports the very minor change in the distribution of the electric field across the slot due to the physical short-circuited connection to the right large ground. On the other hand, this short allows a microstrip line to be placed in a cross configuration below this slot using the metal sheet as ground plane, thus making it possible to arrange circuit components both on the interior and the exterior metal surface of the outer slot.

The inner antenna is a broadband annular slot with a cross-



(a) (b)

Fig. 4. Electric field distribution at 900 MHz for the outer slot (a) without and (b) with a short-circuit connection to the right large ground.

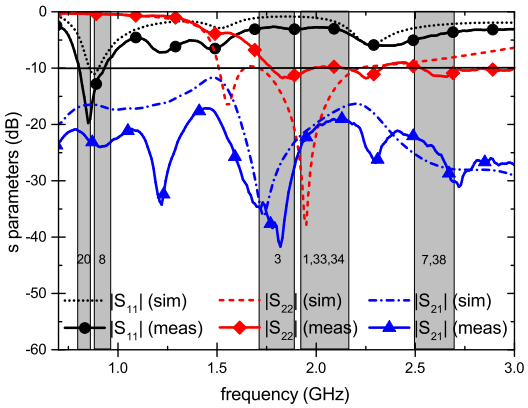


Fig. 5. Scattering parameters of the proposed antenna topology: simulations (solid) and measurements (dashed line). The bands of interest are marked with gray stripes and their number is reported at the bottom of each stripe.

shaped feed line [29]. It consists of a circular slot, a circular patch radiator, which is offset with respect to the slot center, and a proximity-coupled cross-shaped feed line, which also in this case is optimized to obtain a good impedance matching to $50-\Omega$ throughout the whole band of interest ($1.71-2.69$ GHz). All relevant parameters of the antenna are reported in Fig. 3.

Fig. 5 shows a comparison between the simulated and measured scattering parameters of the proposed antenna topology. Number "1" refers to the outer slot, while number "2" is linked to the inner slot, according to the port numeration reported in Fig. 3. The bands involved in the design are marked with gray rectangles. A good matching can be noticed in all bands with the exception of the frequency range on the right part of band 8 and on the left part of band 3. The slight mismatch is on the order of 1-2 dB. Slight discrepancies between simulations and measurements are due to fabrication tolerances. Moreover, the mutual coupling is below -19 dB in all frequencies of interest.

Finally, an estimation of the maximum realized gain of both antennas (linear polarization), which occurs in broadside direction, has been reported (see Fig. 6). The first antenna features a gain of 1.3 dBi at 900 MHz and its maximum gain around 2.3 dBi is achieved at 940 MHz, whereas the second antenna features a gain varying between 4 and 6 dBi in the band of interest. Antenna #1 features a half-power beamwidth (HPBW) of around 108° in the xy plane and 81° in the xz plane at 900 MHz, whereas antenna #2 shows HPBW of 78° and 78.5° in the xy plane, and 71° and 63° in the xz plane at 1.8 and 2.6 GHz, respectively.

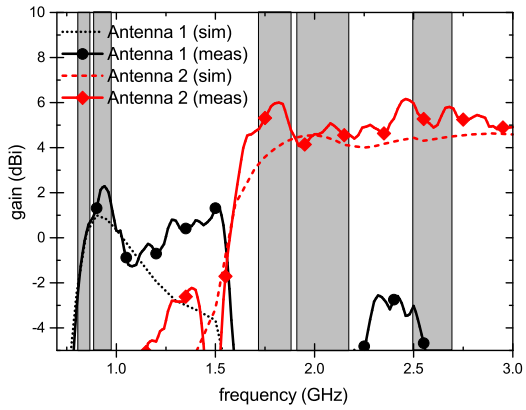


Fig. 6. Measured and simulated gain of the proposed antenna topology.

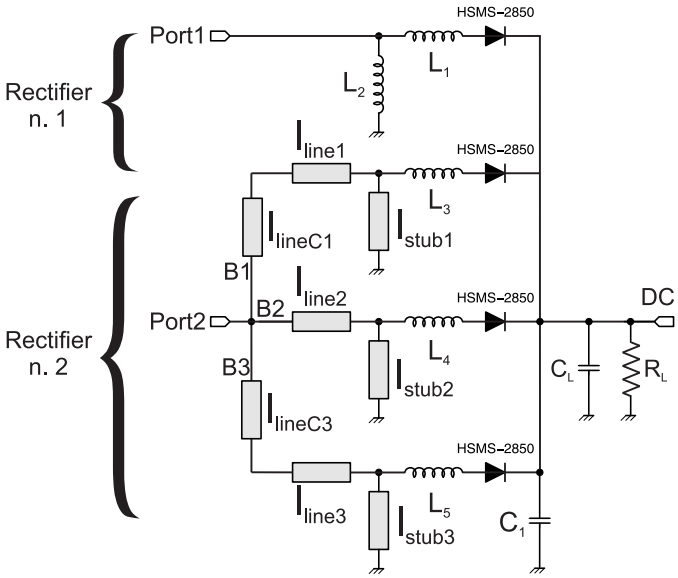


Fig. 7. Schematic of the multiband rectifier. The main circuit parameters are: $L_1 = 47$ nH, $L_2 = 27$ nH, $L_3 = 10$ nH, $L_4 = 8.7$ nH, $L_5 = 4.7$ nH, $l_{stub1} = 9.1$ mm, $l_{stub2} = 6.9$ mm, $l_{stub3} = 5.9$ mm, $w_{stub} = 0.4$ mm, $l_{lineC1} = 6.5$ mm, $l_{line1} = 11$ mm, $l_{line2} = 13$ mm, $l_{lineC3} = 4.25$ mm, $l_{line3} = 11$ mm, $w_{line} = 1$ mm, $C_1 = 100$ pF, $C_L = 100$ pF and $R_L = 3\,\text{k}\Omega$.

V. Rectifier Design

The design of the rectifier has been carried out with the aim of achieving a good balance among circuit complexity, number of utilized surface mount components, and low-power performance, to be consistent with the inexpensive nature of the adopted substrate.

The schematic of the proposed rectifier is shown in Fig. 7. The circuit parameters have been optimized to achieve the maximum RF-to-dc conversion efficiency for available input power levels ranged between -20 and -15 dBm. The design has been carried out by means of the Advance Design System suite.

As anticipated in Fig. 1, each antenna is connected to its own rectifier. each antenna is connected to its own rectifier. Due to its relatively small bandwidth (19.3 % around 880 MHz) the first rectifier ("rectifier #1" in Fig. 1) was realized with a simple envelope detector. On the other hand, this configuration was impractical for the second rectifier ("rectifier #2"). In fact, because of the frequencydependent input impedance of the diode and the limited bandwidth of common compact matching networks, it was very hard to achieve a reasonable large-signal input matching all over the band (45 % around 2.2 GHz), which is essential to guarantee the maximum power transfer to the load. Examples of multiband single-branch rectifiers have been provided in the literature (see [14], [17], [30], [31]), but such approaches are not able to guarantee the coverage over the intended broad frequency band, while allowing for low-power operation. So, in order to achieve a high RF-to-dc conversion efficiency of the rectifier throughout the band, it has been chosen to connect three envelope detectors in parallel, in such a way that each of them operates in a narrower band, according to the approach reported in [12], effectively synthesizing the aggregate band. Each branch, labeled with "B1", "B2" and "B3" in Fig. 7, has been tuned around 1.8, 2.05 and 2.6 GHz, respectively. By optimizing each branch of the rectifier, it is possible to finetune its conversion efficiency so as to achieve a local maximum in each subband of interest. To make the optimization process as accurate as possible, the S-parameter files provided for the lumped inductors by the manufacturer were introduced in the Advanced Design System schematic (the selected inductors belong to the Coilcraft 0603HP series).

Each utilized envelope detector is based on a seriesconnected HSMS2850 low-barrier Schottky diode from Avago, which is suitable for operation in the μW range [32]. Each diode is connected to an input matching network, consisting of an L-type impedance transformer, aimed at converting the input impedance of the antenna (around 50 Ω) to the conjugate of the large-signal input impedance of the diode (see [33]). Unlike in [19], here, some transmission line sections have been replaced by lumped (Coilcraft) components. In fact, it has been observed that the utilized rectifier topology is particularly sensitive to the losses of the line which connects the diode to the stub, since in this section, there is a higher voltage drop (this is actually the aim of the transformer). It was observed that the Coilcraft lumped components are less lossy for the inductance levels of interest in the circuit, as shown in Fig. 8. Simulations were conducted, considering both the S-parameter files of the inductors provided by Coilcraft and the high-impedance microstrip lines connected to ground. The technological pitch of 0.2 mm was chosen for the linewidth. The simulations were performed for both the smallest and the highest inductance values utilized in the rectifier, and in both cases the superiority of the Coilcraft components is noticeable, although this effect is more prominent for the 47-nH inductance. In the first rectifier, the stub has been replaced as well by an inductance due to its low frequency of operation, to reduce the circuit size.

In addition to "lineC1" and "lineC3", introduced for the sake of feasibility, three transmission line sections ("line1", "line2" and "line3" in Fig. 7) have been added to the three branches of the second rectifier in order to rotate the input impedance of each branch, so that the frequencies outside the operating subband feature a high impedance. On the contrary, the operating center frequency of each branch is only marginally affected by its respective line section, since the circuit is designed in such a way that the input impedance at this frequency is approximately located at the center of the Smith chart.

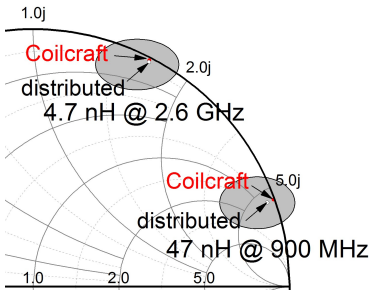


Fig. 8. Inductance implementation: study of the losses related to lumped (Coilcraft) and distributed inductors on paper substrate.

The main idea is to allow each frequency to enter predominantly the branch which is optimized to convert it, thus reducing the consequent loss of power. Fig. 9 reports the input reflection coefficient of each branch of the second rectifier with and without the respective input transmission line sections ("line1", "line2" and "line3" are mentioned as "input line" in Fig. 9), whereas Fig. 10 shows the transmission coefficient between Port2 and each branch of the second rectifier. All reported results were calculated for the design power level of 15 dBm. The addition of these lines was shown to improve the transmission coefficient by up to 4 dB for each branch of the rectifier and for all frequencies of interest. This approach is analogous to that applied to frequency division multiplexers and, although suboptimal for branches B1 and B3 (only one external band can be rotated to open), it eliminates the need to implement high-order bandpass filters [12].

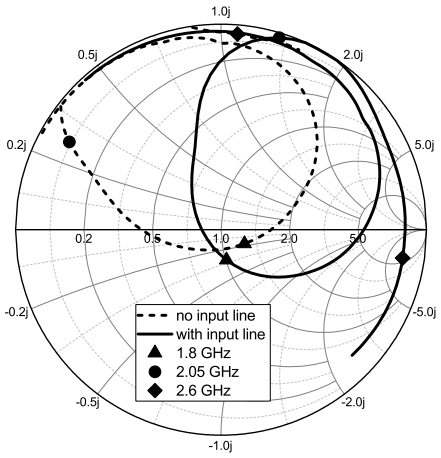
Finally, all branches were connected to a single RC filter, with the exception of the last branch that was tuned to 2.6 GHz, where an additional capacitor was introduced in order to minimize interferences among detectors. The load, equal to 3 kΩ, was a parameter of the rectifier optimization (see Fig. 17).

In the end, the whole multiband on-paper rectifier was realized with four diodes, five inductors, two capacitors, and the load.

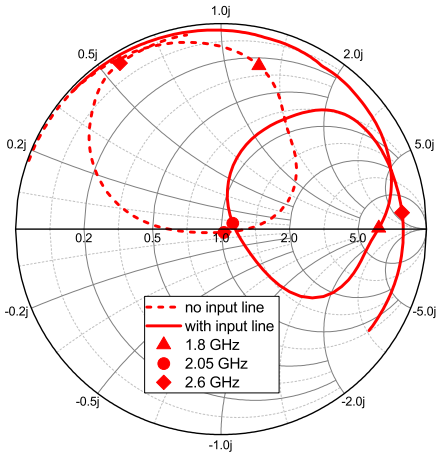
A photograph of the rectifier prototype is shown in Fig. 11.

After a fine-tuning of the value of the lumped components, which has been conducted in order to compensate for additional parasitics not included in the simulations (such as the soldering effects on the paper and the glue of the copper tape, the connectors impact, and so on), the performance of the rectifier under a single tone sweep has been evaluated for different power levels, and the results are reported in Figs. 12-19.

Fig. 12 shows the adequate large signal input reflection coefficient at both ports of the rectifier for a -15-dBm avail-



(a)



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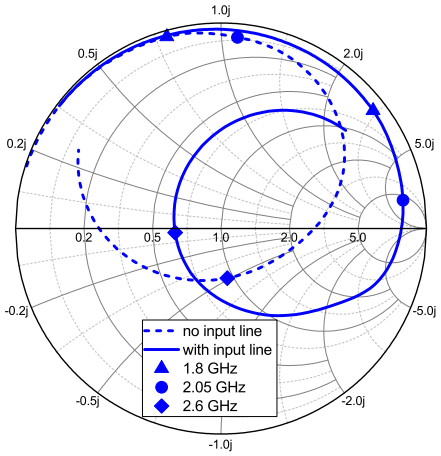


(c)

Fig. 9. Study of rectifier n.2: input reflection coefficient of each branch [(a) B1, (b) B2, and (c) B3] with and without transmission line sections. The simulation results are reported for an available input power of 15 dBm.

able input power. An $|S_{11}|$ of less that -6 dB was achieved throughout all bands of interest.

In Fig. 13 the rectifier performance in terms of output dc voltage is demonstrated, whereas Fig. 14 shows the resulting

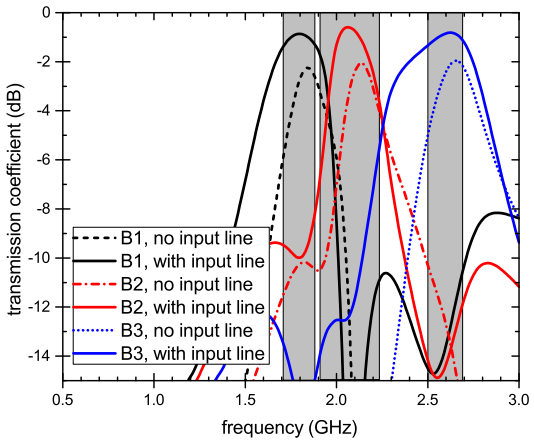


Fig. 10. Transmission coefficient for each branch of rectifier n.2 with and without transmission line sections. The simulation results are reported for an available input power of -15 dBm.

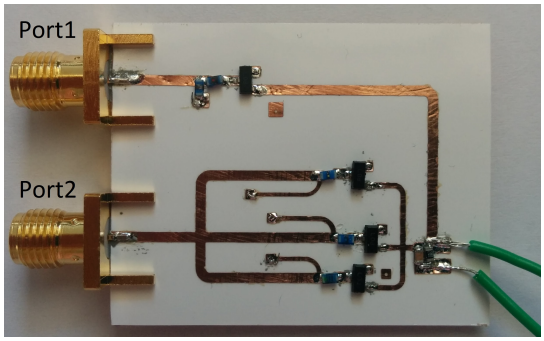


Fig. 11. Photograph of the rectifier prototype. Active area: $3.5 \times 3\,\mathrm{cm}^2$.

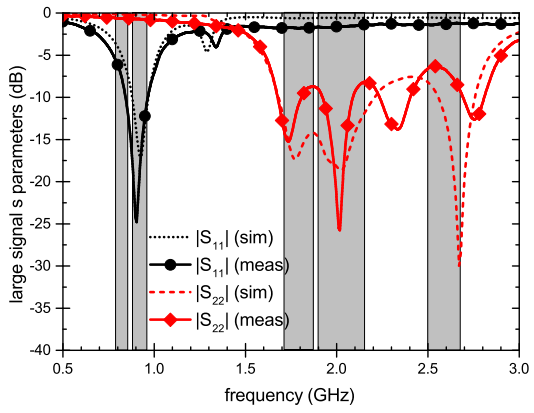
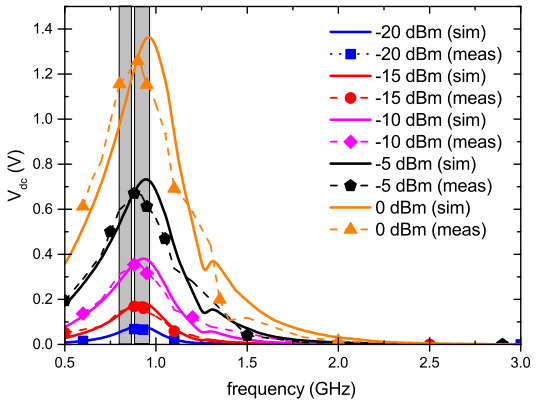


Fig. 12. Rectifier input reflection coefficient for -15-dBm available input power: comparison between simulations and measurements.

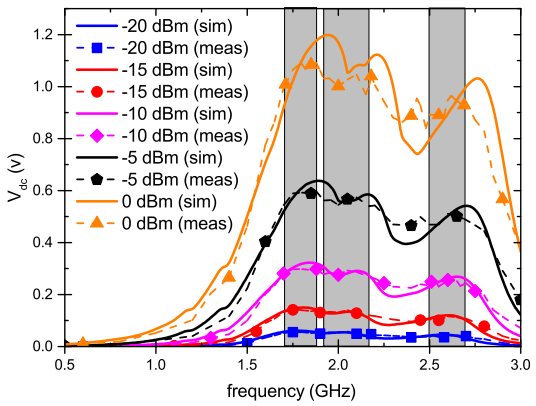
RF-to-dc conversion efficiency, which has been derived from the measured voltage according to

$$\eta_{RF-dc} = \frac{V_{dc}^2/R_L}{P_{avs}} \times 100, \tag{1}$$

where R_L is the load resistance, V_{dc}^2/R_L represents the dc

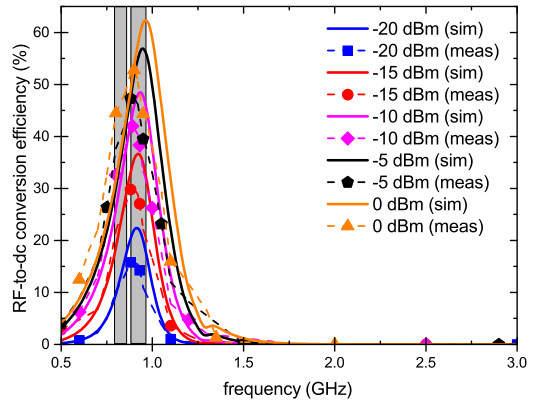


(a)

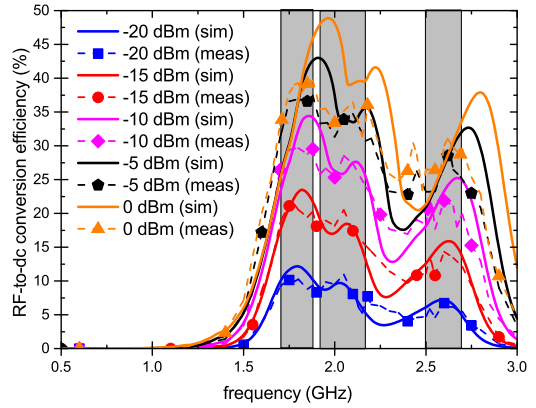


(b)

Fig. 13. Multiband rectifier: output dc voltage for a single tone sweep. (a) Port1. (b) Port2.



(a)



(b)

Fig. 14. Multi-band rectifier: RF-to-dc conversion efficiency for a single tone sweep. (a) Port1 and (b) Port2.

power delivered to the load, and P_{avs} is the available RF power at the source. The measurement has been separately carried out for Port1 [reported in Fig. 14(a)] and for Port2 [in Fig. 14(b)].

The first rectifier, which features the best performance, shows an output dc voltage in the operating band ranging from 50 to 70 mV and from 140 to 170 mV for the available input power levels of 20 and 15 dBm, respectively, corresponding to peak efficiencies of $16\,\%$ and $30\,\%$. Moreover, its peak efficiency grows up to the $52\,\%$ at 0 dBm.

On the other hand, the second rectifier presents an output dc voltage in the range of 36-57 mV and 101-142 mV at -20 and -15 dBm, with the respective peak efficiencies of $10.2\,\%$ and $21.3\,\%$, which are achieved in the first band (band 3 of Table I). In the other two bands, the respective peak efficiencies of $11\,\%$ and $6\,\%$ are obtained for -20 dBm and of $20.5\,\%$ and $14.3\,\%$ for -15 dBm. This performance deterioration is mainly due to the diode losses, which increase with frequency [32], and to the input interaction and interference among the three branches, which is missing, instead, in the first rectifier.

However, it is worth noticing that the circuit performance varies quite smoothly in all bands for all power levels under test, thus achieving the goal to ensure energy harvesting from all bands of interest.

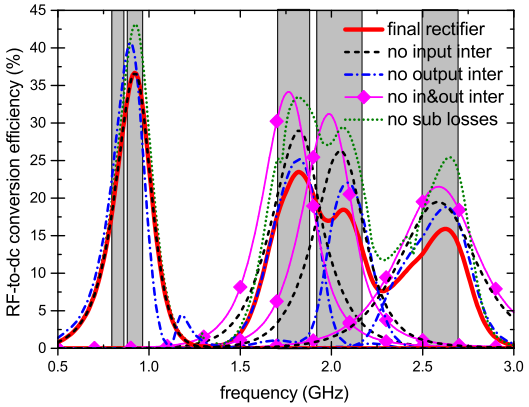


Fig. 15. Multiband rectifier: simulation study of the impact of substrate losses, input and output interactions.

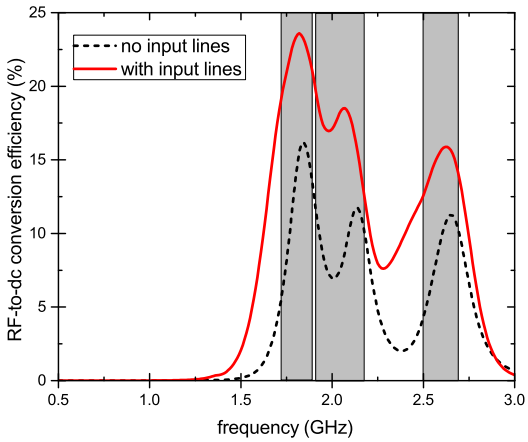


Fig. 16. Simulated RF-to-dc conversion efficiency of rectifier #2 with and without input transmission line sections for an available input power of -15 dBm.

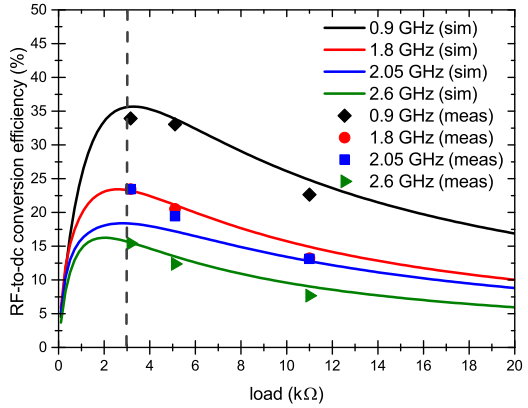


Fig. 17. Multiband rectifier: RF-to-dc conversion efficiency versus load (P_{avs}=-15 dBm).

In Fig. 15, an analysis of the losses associated with both the substrate and the chosen architecture is reported. Four simulations regarding four different scenarios are considered. In the first one, all branches are separated at the input, and the source is connected to one branch at a time, while the others are connected to a $50-\Omega$ impedance, thus removing the effect of input coupling; in the second one, all rectifier branches are connected to a different output RC filter, thus removing their output interaction; then, the performance of each branch of the rectifier is separately evaluated, i.e., each branch is connected to a different source and load (no input and output interactions); finally, in the fourth one, the substrate losses have been removed without changing the final circuit topology. As a case study, the RF-to-dc conversion efficiency for an available input power of 15 dBm is considered. It is apparent that the major limitation to the rectifier efficiency is represented by the substrate losses, which causes an efficiency reduction around $10\,\%$ at all center bands. The aggregate losses related to the circuit topology with both input and output coupling are on the whole comparable with the dielectric ones, with the input and output interactions having an impact on average around $7\,\%$ and $3\,\%$, respectively. Nevertheless, the maximum efficiency achieved in [19] around 2.4 GHz is intermediate between the ones of the rectifiers operating around 2.05 and 2.6 GHz without any input and output interactions. It is worth noticing that the branch centered around 900 MHz is only affected by an output interaction, thus experiencing a lower efficiency degradation. Additionally, Fig. 16 highlights the beneficial impact of the input transmission line sections on the conversion efficiency of the second rectifier due to the more selective bandpass characteristic experienced by the input signal in each branch. The simulations were carried out for the 15-dBm available input power and show an improvement of the peak performance in each subband around $7\,\%$.

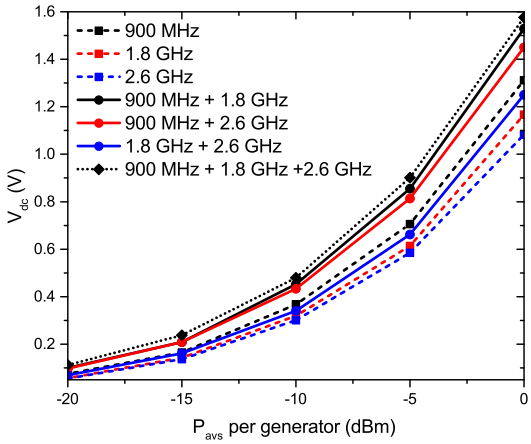


Fig. 18. Multitone test: measured output dc voltage versus an available input power (P_{avs}) per source for one, two and three tones.

Fig. 17 reports the RF-to-dc conversion efficiency with respect to the load resistance at the design power of 15 dBm. The optimum load is different for the four branches of the rectifier, and the value of $3\,\mathrm{k}\Omega$ is chosen as a result of a tradeoff aimed at maximizing the output dc power under the assumption that an equal amount of power enters each branch of the rectifier at its respective central frequency. This is equivalent to maximizing the sum of the efficiencies calculated at the four central frequencies, as reported in

$$\begin{aligned} P_{dc} &= P_{dc,0.9} + P_{dc,1.8} + P_{dc,2.05} + P_{dc,2.6} = \\ &= \eta_{0.9} P_{avs} + \eta_{1.8} P_{avs} + \eta_{2.05} P_{avs} + \eta_{2.6} P_{avs} = \\ &= P_{avs}(\eta_{0.9} + \eta_{1.8} + \eta_{2.05} + \eta_{2.6}), \end{aligned} \quad (2)$$

where P_{dc} is the total output dc power, $P_{dc,x}$ is the output dc power corresponding to an RF input power P_{avs} at a frequency of x GHz, and η_x is the RF-to-dc conversion efficiency associated with the P_{avs} input power and calculated at a frequency of x GHz. Therefore, this approach penalizes the rectifier efficiency at 2.6 GHz, which features a maximum efficiency for a load equal to $2\,\mathrm{k}\Omega$.

Finally, the circuit performance under multitone input signals has been tested. The rectifier has been connected to

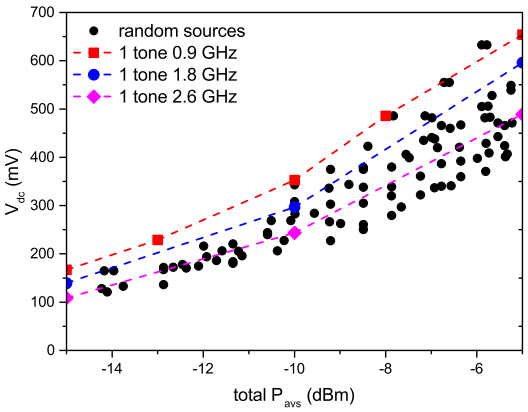
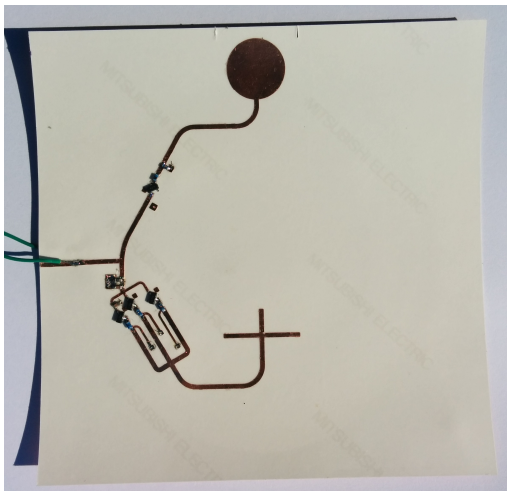


Fig. 19. Multione test: measured output dc voltage for an available RF input power randomly distributed over three tones.

(a)



(b)

Fig. 20. Layout of the rectenna prototype. (a) Top side. (b) Bottom side.

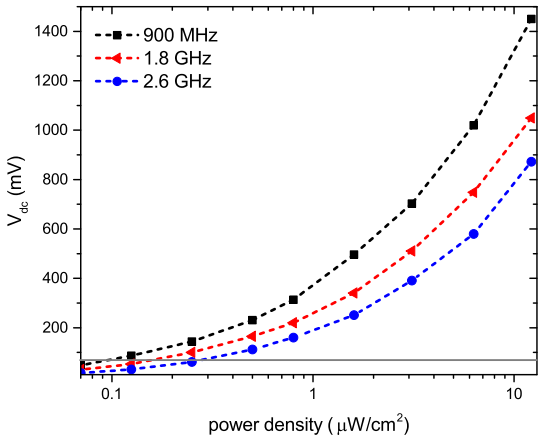
three different power generators, each of them transmitting a single tone tuned to 900 MHz, 1.8 GHz, and 2.6 GHz, respectively. The output dc voltage is recorded as a function of the available input power per source for one, two, and three tones. In this way, the rectifier capability to take advantage of the simultaneous presence of multiple tones has been verified. As a second experiment, the power associated with each tone has been randomly varied in such a way that their sum takes values within the range [15, 5] dBm. This operation has been repeated for 100 times, and the output dc voltage has been recorded with results shown in Fig. 19. It is worth noticing that the data variance is mainly due to the different conversion efficiencies experienced in each band and that, as expected, the best performance is obtained for the lower band. A few points are above the 900-MHz one-tone performance due to the fact that the phases of the three independent generators are randomly combined to achieve an optimized waveform [34].

Fig. 21. Test of the rectenna in laboratory: experimental setup.

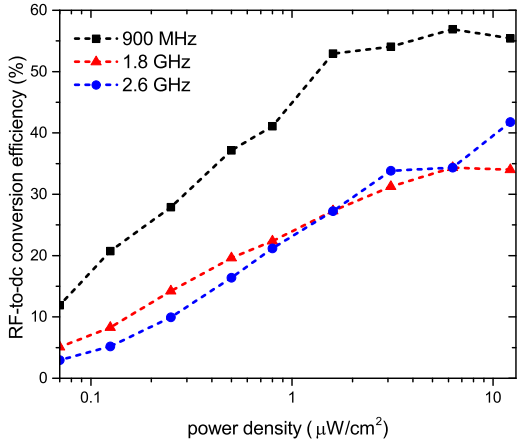
VI. RF Harvester Measurements

In order to validate the proposed architecture, the whole rectenna topology was fabricated on a paper substrate, by using the materials and techniques described in Section III, and tested (see Fig. 20 where both sides of the circuit are shown). The rectenna maximum thickness is equal to 1.2 mm, including both the thickness of the substrate and the lumped components. The two rectifiers were arranged so as to take advantage of the ground plane between the two antennas. This guarantees that the rectifiers are placed near the antenna feed, thus reducing the presence of losses. Moreover, thanks to this arrangement, the rectenna area corresponds to the bare area of the outer antenna, confirming the compactness of the proposed solution, whereas the cut in the outer slot allows for the usage of the outer metal surface to host other circuit components. The proposed rectenna prototype features a weight of only 6 g, while it is conformal and suitable for a broad range of low-cost applications. For instance, it can be used together with a suitable energy storage unit and a circuit with very low duty cycle operation ensuring a large energy autonomy and batteryless operation. Such a system can also be used as a trickle charge for batteries in order to compensate for their self-discharging.

The rectenna performance was first measured in a controlled environment, as reported in Section VI-A-A, and finally tested in real scenarios, both in normal (flat) and in bent conditions.



(a)



(b)

Fig. 22. Test of the rectenna in laboratory. (a) Output dc voltage and (b) RF-to-dc conversion efficiency versus power density. Gray solid line: activation voltage threshold of the state-of-the-art dc/dc converters.

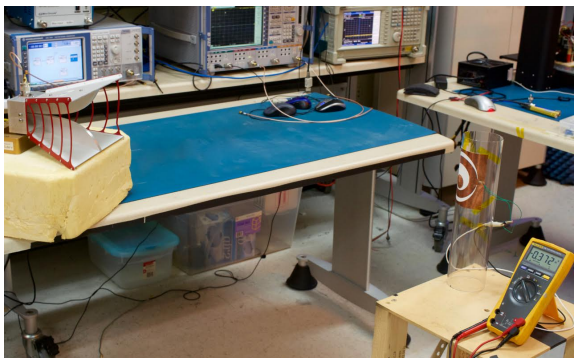


Fig. 23. Rectenna testing under bending conditions: experimental setup.

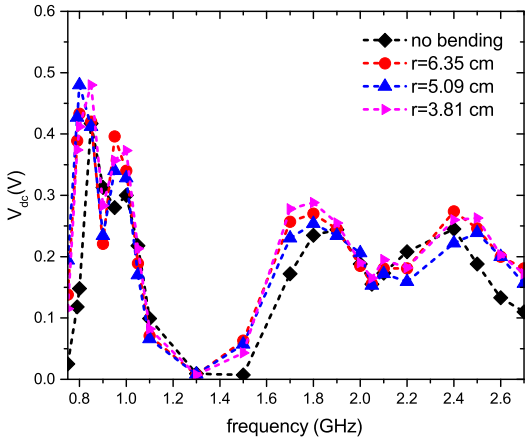


Fig. 24. Rectenna performance under different bending conditions: output dc voltage for different radii of curvature.

A. Rectenna characterization in a controlled environment

A controlled experiment was performed in a laboratory room using the setup shown in Fig. 21. A power source was connected to a double-ridged waveguide horn antenna (3115 model from ETS-LINDGREN). The rectenna was aligned with the antenna and placed at a distance of 1.6 m, which is outside the Fraunhofer region, delimited by the well-known equation (3), for both the transmitting and the receiving antenna

$$R = \frac{2\,D^2}{\lambda_0}, \tag{3}$$

where R is the distance between the phase center of the antennas, D is the largest dimension of the power-source antenna (equal to 25 cm for the horn), and λ_0 is the free-space wavelength associated with the operating frequency.

The rectenna was connected to a multimeter, and the power source was set to three different frequencies (900 MHz, 1.8 GHz, and 2.6 GHz), while for each tone the transmitted power was varied from 5 to 29 dBm. Fig. 2222(a) and (b) shows the measured output dc voltage and RF-to-dc conversion efficiency as a function of the power density at the antenna of the harvester. To calculate the latter quantity, the "standalone" antenna topology on paper was placed in the same position as the rectenna under test, and it was connected to a power meter. The measured RF input power was divided by the effective area of the respective slot antenna, estimated with the well-known equation as

$$A_{eff} = G_i \frac{{\lambda_0}^2}{4\,\pi}, \tag{4}$$

where G_i is the measured gain for the respective-to-the-band slot antenna. In particular, the power density was calculated for the three frequencies of interest for each power level, and the results were averaged in order to compensate for uncertainties stemming from the experimental setup.

According to the obtained results, a power density in the order of 0.1-1 μW/cm^2 is sufficient to activate emerging ultralow voltage electronics, while 10 μW/cm^2 is required for standard electronics ("switch ON" voltage around and above 1 V).

TABLE II
COMPARISON WITH STATE-OF-THE-ART RF ENERGY HARVESTERS

Ref.	Freq. (GHz)	power density (μW/cm^2)	Max Eff. (%)	V_{dc} (V)	Area (cm^2)	Substr.	tanδ	Tech.
this work	**0.79-0.96; 1.71-2.69**	**6**	**57**	**1.02**	**11x11**	paper	0.05	copper tape
[19]	2.4-2.5	3	28	0.26	4.9x5.4	paper	0.05	copper tape
[35]	0.8-0.9; 1.8-1.9	0.2	15	/	13x13	PET	0.08	laser etching
[36]	0.9	8	48.6	/	15.5x5.4	PET	/	inkjet
[37]	2.45	3	1.6	0.2	6x4.5	cardboard	0.042	inkjet
[38]	0.86-0.918	14	50	1.42	19x13.9	pile/jeans	0.009/0.023	conductive non-woven fabric
[39]	0.9; 1.75; 2.45	0.7	19	0.13	17x17	pile/Kapton	-/0.002	Global EMC shielding fabric

B. Rectenna testing in real scenarios

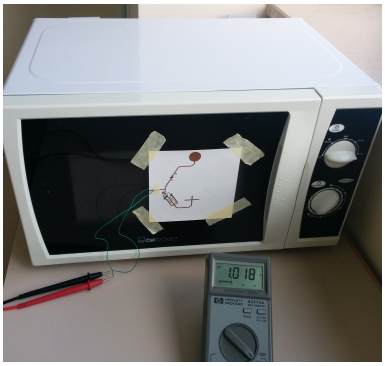
The performance of the presented flexible rectenna was tested under different bending conditions (see Fig. 23). The rectenna was placed at a distance of 90 cm and aligned with the transmitting antenna. The latter was connected to a signal generator set to a fixed transmitting power of 15 dBm. The recorded output dc voltages versus frequency were collected for three radii of curvature, as shown in Fig. 24. The impact of bending seems to be neglectable, and the rectenna performance remains nearly unchanged for a radius of curvature as small as 3.81 cm.

The proposed rectenna was proved to be suitable for utilization in a broad range of realistic scenarios, with numerous ambient devices using microwaves for different purposes, thus effectively increasing the local RF power density. For instance, the rectenna was mounted on the door of a microwave oven, which operates at the ISM S-band [Fig. 25(a)]. Despite the power shielding, an output dc voltage around 1 V was measured across the 3-kΩ load during the whole cooking time. In the second example, the rectenna was placed under a smartphone. During an incoming call (GSM900 band), an output dc voltage around 300400 mV was observed.

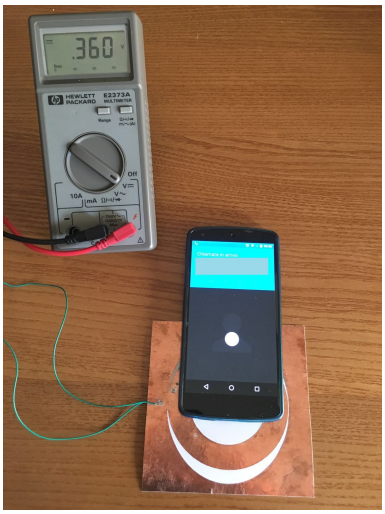
C. Performance Comparison

In Table II, a comparison of the presented rectenna with other state-of-the-art rectennas fabricated on flexible and/or unconventional substrates is reported. Reference [19] has already been presented in the introduction, and consists of a single annular slot and an envelope detector. A dual-band rectifier and a single wideband monopole were used in [35], while [36] was based on a class-F rectifier and a direct matching between the antenna and the diode. In [37] a half-wave double monopole antenna and a two diode rectifier were employed, and finally, multilayer wearable rectennas were proposed both in [38] and [39]. Reference [38] relies on a rectangular patch antenna, while [39] is based on a slot-loaded annular ring antenna; both solutions utilize a single-stage full-wave rectifier. As it can be easily observed, the majority of these papers focus on single-band or even single-frequency harvesters.

The presented rectenna features a very good performance in terms of occupied area and maximum efficiency with respect to incident power density, while it guarantees operation over a broader frequency range over multiple bands.



(a)



(b)

Fig. 25. Photograph of possible practical scenarios for the proposed rectenna: ambient RF energy harvesting from (a) operating microwave oven and (b) cell phone during an incoming call.

VII. CONCLUSION

A novel compact multiband rectenna on paper for ambient RF energy harvesting has been introduced in this effort. Both the design procedure and the measurement results have been thoroughly discussed. The rectenna is able to harvest power in all LTE frequency bands at the available input power levels as low as 20 dBm in a variety of practical and conformal scenarios, thus setting the foundation of truly autonomous IoT and smart node configurations.

Future improvements are also possible: the rectifier can be

optimized for multitone operation, for instance, by means of the adoption of higher order bandpass filters in each branch of the second rectifier [12]; the applicability of the system can be freed from orientation constraints by developing compact broadband circularly polarized antennas, achievable through the introduction of ad hoc shorts in specific positions of the slot antennas [28].

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Antennas and Metamaterials for Electromagnetic Energy Harvesting

by

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I hereby declare that I am the sole author of this thesis. This is a true copy of the thesis, including any required final revisions, as accepted by my examiners.

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Abstract

The emergence of microwave energy harvesting systems, commonly referred to as rectenna or Wireless Power Transfer (WPT) systems, has enabled numerous applications in many areas since their primary goal is to recycle the ambient microwave energy. In such systems, microstrip antennas are used as the main source for collecting the electromagnetic energy.

In this work, a novel collector based on metamaterial particles, in what is known as a Split Ring Resonator (SRR), to harvest electromagnetic energy is presented. Such collectors are much smaller in size and more efficient than existing collectors (antennas). A feasibility study of SRRs to harvest electromagnetic energy is conducted using a full wave simulator (HFSS). To prove the concept, a 5.8 GHz SRR is designed and fabricated and then tested using a power source, an Infiniium oscilloscope and a commercially available patch antenna array. When excited by a plane wave with an H-field normal to the structure, a voltage build up of 611 mV is measured across a surface mount resistive load inserted in the gap of a single loop SRR. In addition, a new efficiency concept is introduced, taking into account the microwave-to-AC conversion efficiency which is missing from earlier work. Finally, a 9X9 SRR array is compared with a 2X2 patch antenna array, both placed in a fixed footprint. The simulation results show that the array of SRRs can harvest electromagnetic energy more efficiently and over a wider bandwidth range.

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Dedication

To my parents

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Chapter 1

Introduction

1.1 Thesis Objective

The work presented in this thesis is aimed at harvesting electromagnetic energy. In the literature, this topic is presented as rectenna systems or Wireless Power Transfer (WPT) systems. Therefore, both systems should be reviewed and the differences between the two systems should be understood in order to contribute to this topic. Hence, there are three main objectives of this thesis:

- Review the history of microwave energy harvesting through the work presented in the literature under rectenna and WPT systems. *(Chapter 2)*
- Analyze the challenges, limitations and working mechanism of rectenna systems through simulation and fabrication. *(Chapter 3)*
- Propose new and novel structures (metamaterial particles) for collecting electromagnetic energy, which are smaller in size and more efficient than existing collectors (antennas). *(Chapter 4)*

1.2 Thesis Contribution

The research work presented in this thesis provides three main contributions: a novel collector based on metamaterial particles for harvesting electromagnetic energy is introduced. A proof of concept through simulation and measurement is provided. In addition, a new method for calculating the efficiency of collectors is proposed to account for the microwave-to-AC conversion efficiency which is missing from earlier work presented in the literature. Finally, a simulation study is conducted which compares an array of patch antennas and an array of metamaterial particles in terms of microwave-to-AC conversion efficiency, with arrays both occupying the same footprint.

1.3 Thesis Outline

The reminder of the thesis is organized as follows:

Chapter 2 reviews the history and development of Wireless Power Transfer (WPT) and rectenna systems. In addition, the working mechanism of WPT and rectenna systems are provided. This chapter also reviews the compromises and challenges one can face when designing a full rectenna system. Finally, the most common applications of rectenna systems and their working mechanisms are covered.

Chapter 3 provides detailed procedures for designing a full rectenna system operating at 2.45 GHz. The steps taken during the fabrication process are also listed. A microstrip-line fed patch antenna is then fabricated and its results were compared to the once obtained by simulation. A strong agreement is observed in the return loss plot. Finally, the rectifier circuit is fabricated and tested.

Chapter 4 proposes a novel and new energy collector based on metamaterials in what is known as a Split Ring Resonator (SRR). A brief overview of metamaterials and some of their applications is presented. Two validity tests are conducted through simulation to study the feasibility of using one type of SRRs to harvest electromagnetic energy. Then an experiment is conducted though testing and measurements; it yielded results that prove the concept of using SRRs as collectors. Finally, the efficiency of an array of 9 X 9 SRRs

is compared to that of an array of 2 X 2 patch antennas placed over an identical footprint. The SRR array resulted in at least 25% more efficiency and a wider bandwidth ,over which the energy is collected, than that of the antenna array.

Chapter 5 finally, reviews the main milestones presented in this thesis. In addition, some concluding remarks and suggestions for future work are provided.

Chapter 2

Background

2.1 Introduction

This chapter provides a comprehensive overview of the most important milestones in the history and development of WPT systems, from the work and experiments of Hertz up to the date of this thesis. In addition, the working mechanism of WPT and rectenna systems, and some existing design challenges are discussed. Finally, a number of promising applications of rectenna systems along with a brief description of their operating mechanism are presented.

2.2 Historical Milestones

2.2.1 Early History

The notion of Wireless Power Transfer (WPT) was sparked in 1888 when Heinrich Hertz demonstrated electromagnetic wave propagation in free space [13]. Hertz used half wavelength dipoles operating at 500 MHz along with parabolic reflectors for transmitting and receiving electromagnetic energy. The high frequency power was generated by charging the opposite halves of the dipoles with high voltage in order to develop an arc across the gap.

As a result energy is stored in the gap, forming oscillations, and then the half wavelength dipoles radiate some of the oscillating energy as microwaves [52]. Hertz's experimental work constitutes crucial validation of Maxwell's hypothesis in regards to electromagnetic propagation in free space.

At the turn of the 20th century, further experiments on the concept of wireless power transfer were conducted by Nickola Tesla. A grant of $30,000, offered by Colonel John Jacob Astor, was used to build a gigantic coil resonating at 150 kHz, which was then placed in a large building in Colorado Springs, Colorado. Over the coil, was raised a 60 m pole topped by a 1 m diameter copper ball. The coil was then excited with low frequency power of 300 kW obtained from the Colorado Springs' electric power company. Consequently, a large RF potential was produced on the sphere, reaching 100 MV. After all his hard work at the Colorado Springs laboratory, Tesla was able to light 200 light bulbs placed 42 km away from the base station [18]. Tesla's work intrigued many researchers in Japan [57] and the US [1];however, technological limitations prevented them from progressing in this area.

It was not until the 1950s that the WPT became a useful technology, with the development of high power microwave tubes by the Raytheon Company. These tubes were considered essential for transmitting efficient high power at GHz frequencies. In 1963, William C. Brown, a pioneer in the concept and development of WPT, and his colleagues, demonstrated the first complete and efficient WPT system to a group of United States Air Force officials who funded the high power microwave tubes. In the demonstration, DC power was first converted to microwaves at 2.45 GHz and a power level of 400 W. A horn antenna was used to transmit the generated microwave energy, illuminating a reflector, which then bounced back the energy to its focal point. At the focal point, a diagonal horn antenna was placed to collect the microwave energy, which was converted back to DC by means of thermionic diodes. Conversion efficiencies of only 50% were realized [52].

Due to the low AC to DC conversion efficiency, which is attributed mainly to the diode, researchers have since focused on developing new and more efficient diodes. As a result, Hewlett-Packard Associates offered their new silicon Schottky-barrier diodes to C. Brown's group [12]. This type, the 2900 diodes, has the following advantages: much greater efficiency, much smaller size, and better capability of handling power than any of the previous diodes.

Both the development of high power microwave tubes and of the silicon Schottky-barrier diodes represent a paradigm shift in the history and development of the WPT technology.

In 1975, another key experiment was conducted in the Mojave Desert at the Jet Propulsion Laboratory's (JPL) Goldstone Facility to show the feasibility of transferring microwave power over a long distance. In this experiment, a new record was reached in terms of the amount of power received and the total distance over which the power was transmitted. The setup of the experiment consisted of a high powered klystron, which created microwave energy at 2.388 GHz and then fed it to a 26-m transmitting parabolic antenna as shown in Figure 2.1 [52]. This antenna excited a 26 m^2 rectenna (rectifying antenna) array located 1.54 km away. The rectenna array contained 4590 rectenna elements arranged in 17 panels, each panel consisting of 270 rectennas. Each rectenna element utilized a dipole antenna 0.47λ in length for capturing the microwave energy. Every two adjacent dipole antennas were separated by 0.6λ to maximize the collected microwave energy [19]. A set of lights were used as loads, each rectenna panel devoted to powering two lights. The total DC power collected by the rectenna array was over 30 kW, representing a total conversion efficiency of 84%. Overall, the developed WPT system in the Goldstone experiment was deemed successful for its ability to transmit power over a long distance, to demonstrate high conversion efficiency and to survive various weather conditions.

It is of interest to note that in parallel to the work accomplished by the Americans in regards to WPT, Canadian and Japanese researchers have developed their own programs, which contributed significantly to the advancement of WPT. An example of such a program developed by Canadian researchers is the Stationary High Altitude Relay Program (SHARP) and for those conducted by the Japanese are the Microwave Ionosphere Nonlinear Interaction Experiment (MINIX) and the Microwave Lifted Airplane Experiment (MILAX). For brevity, those programs are only mentioned ; however, details of each program can be found in [46, 33, 24] respectively.

Since the total efficiency of the WPT system depends significantly on the receiving system, commonly referred to as the rectenna system, research on this subject has focused on developing various rectenna systems that yield higher conversion efficiencies. In 1977, W. C. Brown at Raytheon developed a 2.45 GHz rectenna system that reached a conversion efficiency of 90.6% [11]. Both the dipole antenna and the transmission lines used in this

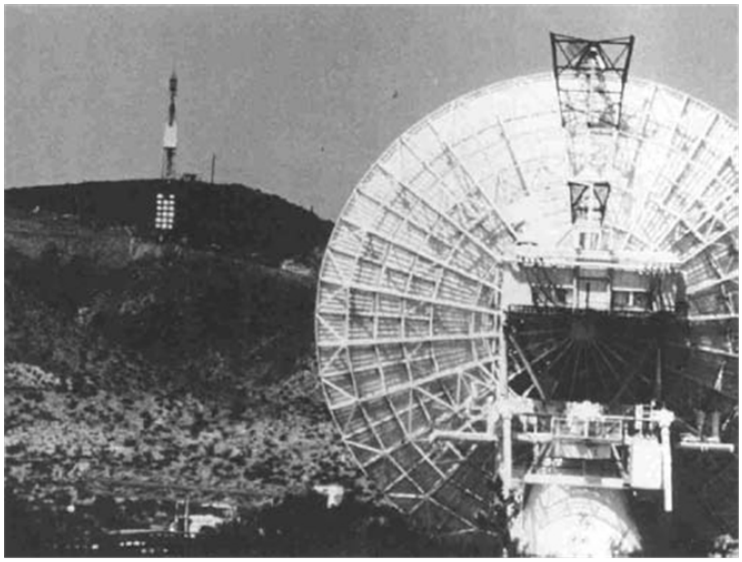


Figure 2.1: A picture showing the experiment conducted at the JPL Goldstone facility in the Mojave Desert.

system were made out of aluminum bars, and the microwave power collected by the antenna was converted to DC using a GaAs-Pt Schottky barrier diode. The conversion efficiency of this rectenna system is recognized as one of the greatest efficiencies ever recorded [34].

In 1985, Brown and Triner developed a revolutionary rectenna system using printed thin film technology. The rectenna system was designed to operate at 2.45 GHz and was able to achieve 85% conversion efficiency [14]. The main advantage of using such technology is its light weight, being 10 times lighter than previously developed rectennas. Since then, most rectenna systems have been made using photolithographic etching techniques for their ease of fabrication, low cost and light weight [52].

2.2.2 Recent Development

In most of the aforementioned experiments, rectenna systems have been designed to operate at around 2.45 GHz due to the limitations of the microwave energy sources available then. However, in 1991 a 35 GHz rectenna was developed by ARCO Power Technologies, with a

total conversion efficiency of 72% [30]. A low atmospheric absorption is the main advantage of operating in the vicinity of this frequency but doing so requires the use of more expensive and inefficient components [34].

In 1992, to compromise among efficiency, component cost and size, the first C band rectenna array was designed, operating at 5.87 GHz. The rectenna array was comprised of 1000 dipole antennas, which fed rectifier circuits with a silicon quad-bridge Schottky diode having a high reverse voltage. This new diode has the advantage of higher power-handling capabilities than those of the previously used GaAs diodes [10]. Conversion efficiencies exceeding 80% were obtained.

Up to that time, the main tools used to analyze and design rectennas were Libra [60], Touchstone [2], and Transmission Line Models [10]. It was not until 1997 that a full wave simulator was used by K. Chang's group at Texas A&M University to design a complete rectenna system. They have used I3ED by Zeland to iteratively design the length and width of the microstrip-lines of a dipole antenna, a low pass filter and the transmission lines linking them. A total RF-to-DC conversion efficiency of 82% was achieved with input power levels of 50 mW at 5.8 GHz [35]. After this point, most of the designs of rectenna systems have incorporated highly efficient full wave simulators such as HFSS, ADS, I3ED and CST. Using such advanced programs allows the designer to simulate both the antenna and the passive components simultaneously.

For the last decade, researchers have focused on improving different aspects of rectenna systems to suit the targeted applications. Such aspects include, but are not limited to:

- Antenna polarization (*i.e.,* linear/circular)
- Antenna type (*i.e.,* rectangular patch, bowtie, dipole, etc.)
- Operating frequency (*i.e.,* single/dual)
- Total size of the rectenna system
- Type of filter (*i.e.,* low-pass/band-pass)
- Impedance bandwidth

- Type of diode
- Rectifier circuit (*i.e.,* half-wave, full-wave, voltage doubler, etc.)
- Substrate material
- Feeding mechanism (*i.e.,* probe, micro-strip, aperture, etc.)

There is no absolute freedom in optimizing each aspect independently as a change in one aspect will alter one or more other aspects undesirably. Therefore, there is always a trade off when one or more aspects are optimized. Depending on the targeted application, one can balance those aspects to arrive at the desired rectenna performance.

Most of the recent microwave rectenna systems have been designed to operate in the range of frequencies between 1-35 GHz. Total conversion efficiencies of 50-80% are reachable when operating at frequencies in the C-band (*i.e.,* 5.8 GHz) as in [16, 50, 38]. However, the conversion efficiency degrades as the operating frequency increases, as in the case of rectennas working in the X-band [37] and the Ka-band [43], where efficiencies of only 21% and 35% are realized, respectively. Even though operating at higher frequencies exhibits a noticeable degradation in the total conversion efficiency, it has the advantage of miniaturizing the total size of the rectenna system, which is crucial for applications where portability is the primary concern. Hence, there is a trade off between efficiency and size when rectennas are designed at various frequencies.

Since the power availability is not always guaranteed when working at a narrow band of frequencies, operating at dual frequencies increases the reliability of rectenna systems. In such designs, rectennas utilize dual antennas that ,when designed properly, can receive the microwave energy at two ranges of frequencies (*i.e.,* 2.45 and 5.8 GHz) as in [42, 53, 27]. However, using dual-frequency rectennas increases the total size of the system due to the use of an additional antenna. A decrease in the total conversion efficiency is also noticeable since the matching network, the output load, and the diode input impedance must be optimized to maximize the power at both frequencies and, therefore, a compromise is always present between the two operating frequencies. Another practical way to increase the reliability of rectenna systems is by the use of wide band antennas. However, designing

a wide band matching network between the antenna and the microwave diode that always ensures maximum power transfer is somehow an intricate task due to the nonlinear behaviour of the diode, which has an input impedance that varies with the input power level and the operating frequency. Energy harvesting is an example of a rectenna application where power availability at the connected load or device is an important requirement.

Recent work on rectenna systems has incorporated Split Ring Resonators (SRRs) to miniaturize the size of the antenna, as in [61], and to suppress harmonics generated by the rectifier circuit, as in [25]. Ziolkowski and his group at the University of Arizona have designed an electrically small rectenna that operates at 1.5754 GHz, utilizing a metamaterial inspired structure. Using such a structure enables a dramatic reduction in the total size of the rectenna system while operating at lower frequencies, solving the problem faced in previous rectennas [62]. A detailed description of the working mechanism of SRRs and the feasibility of using them to harvest microwave energy is presented in Chapter 4.

2.3 Applications of Rectenna Systems

The advent of rectenna systems has greatly advanced many technologies that strive for energy, from powering an implantable artificial heart to tracking and identifying objects in security systems. Rectennas have countless potential applications since their primary goal is to deliver wireless DC power efficiently. Initially, rectennas were developed by Raytheon Company in response to the American Air Force requiring a device that could be attached to a helicopter and rectify microwave energy when a beam is pointed at the device. Brown and his colleagues developed a rectenna array that contains several thousands of point contact silicon diodes. The array produced a DC power level of approximately 200 W. As a result, the group were able to demonstrate a microwave-powered helicopter that can run for at least 10 hours continuously at an altitude of 28 meters above the transmitting antenna [12].

As a consequence of the success achieved in the helicopter demonstration, researchers showed a great interest in using rectennas to solve the global energy crisis by providing a clean and affordable energy source instead of using the harmful method of burning fossil

fuels. In 1968, Petter E. Glasser proposed a concept that uses rectennas to harvest energy from space in what is known by Space Solar Power (SSP). The working mechanism of the SSP concept can be understood by the following four main steps [22]:

1. Collect the solar energy from space and convert it to electricity;
2. Convert the collected power to microwaves and transmit it to Earth;
3. Receive the microwave energy and convert it back to electricity using large rectenna arrays;
4. Provide the collected energy to the power grid.

An example of rectenna array connected to the power grid is shown in Figure 2.2. The United States Department of Energy (DoE) and the National Aeronautics and Space Administration (NASA) has dedicated extensive research to examining the feasibility of such system. Results have shown a great promise for the future of energy. However, it could take decades before the SSP concept is implemented all over the world since it must outweighs the existing methods in terms of cost, reliability and sustainability.

Another vital application of rectenna systems that has excelled rapidly in the last decade is Radio Frequency Identification Systems (RFIDs). In an operational RFID system, there are two main components: a tag (transponder), which contains electronically stored data, and a reader (interrogator) that can read the stored data of the tag. First, the reader transmits RF energy to the tag; then the tag, which is attached to the object to be identified or tracked, receives the energy and sends back a coded signal to the reader. The reader then retrieves the data and responds accordingly (*i.e.,* activates an alarm) [18]. Here, the tag represents a rectenna system since it mainly consists of an antenna, a diode and a microchip. The tag can be either active or passive depending on the requirement of an internal power supply. An active tag, which requires a secondary source of energy in addition to the rectified microwaves, can read data in the range of between 20-100 meters from the interrogator. A passive tag, on the other hand, requires no internal source of energy and usually can respond to a reader at a shorter range than that of an active tag [58]. RFIDs are commonly used in industrial and commercial companies

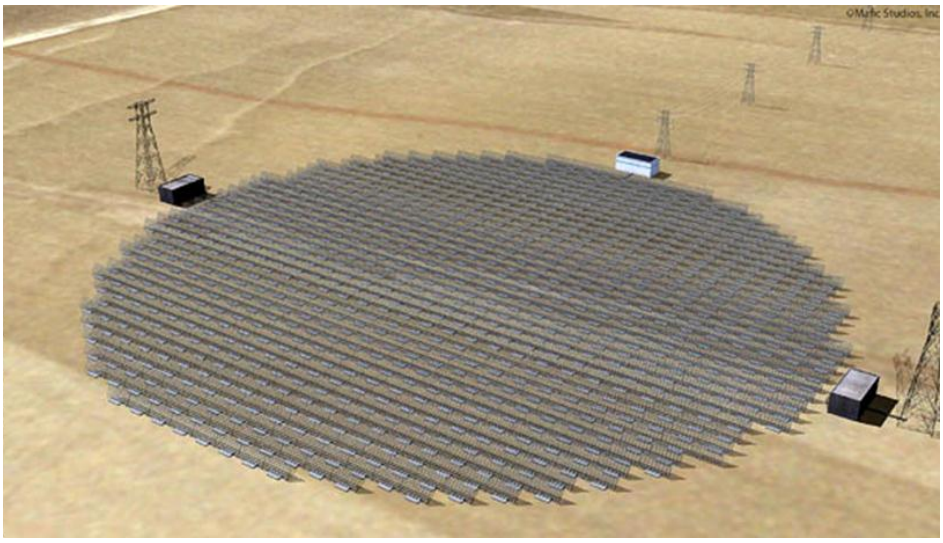


Figure 2.2: A schematic of a receiving rectenna array station connected to the grid. from:http://www.solarfeeds.com/powersat-a-new-space-solar-player/

as tracking and identifying tools for numerous objects, such as for tracking baggage in airports [3], surveillance and security systems [56], tracking pets or endangered animals [29], anti-collision systems [26], and even monitoring oil drill-pipes [51].

Other active research areas where rectennas find promising future applications include but are not limited to driving mechanical actuators at high voltages (50 V) [21], wireless power transfer to a sensor buried in concrete [48], and implantable rectennas for biotelemetry communications [28].

2.4 Operation of Rectenna Systems

A rectenna or a rectifying antenna is part of a larger system known as a wireless power transfer system (WPT). A common WPT system consists of three main subsystems: a transmitter, a travelling medium (*i.e.,* free space), and a receiver, as shown in Figure 2.3. The function of the first block, which represents a transmitter, is to convert DC power to

microwaves using a transmitting circuit and a transmitting antenna. These microwaves will radiate through the transmitting antenna and travel across free space towards a receiver represented by the third block. The receiver, which represents a rectenna system, converts the microwaves to AC power using a receiving antenna. Then the captured AC power is converted back to DC using a rectifying circuit. Hence, the name rectenna, which combines the words of the main components that comprise it. The total efficiency of the WPT system is the product of the transfer function of each functional block, which can be expressed as [40]:

$$\eta_o = \eta_t \times \eta_f \times \eta_r \tag{2.1}$$

where η_o represents the overall efficiency of the WPT system, η_f is the free space efficiency, and η_r denotes the total efficiency of the rectenna system. Usually in WPT systems, the transmitter efficiency η_t is assumed to be unity.

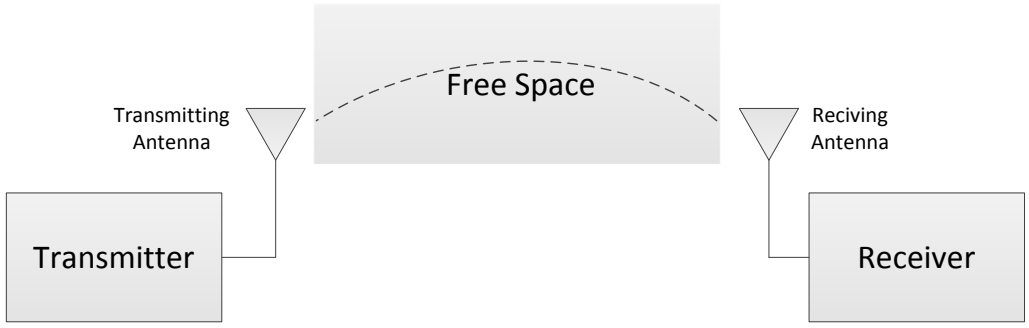


Figure 2.3: Functional blocks of a typical Wireless Power Transfer (WPT) system.

Zooming into the the third block, there are five main elements that form a rectenna system as shown in Figure 2.4. The electromagnetic energy is captured using a receiving antenna operating at the desired frequency. Then, an RF filter is used to transform the antenna impedance to that of the diode, reducing reflections caused by any impedance mismatch. The RF filter also suppresses the unwanted harmonics caused be the nonlinear

behaviour of the diode. After the AC power has channeled maximally from the antenna through the RF filter, a diode (*i.e.,* schottky diode) is used to rectify or convert the collected AC power to DC. To ensure that AC power does not reach the load, a DC filter is inserted after the diode which mainly consists of a shunt capacitor that acts as a short circuit for higher frequencies. Then, a resistive load is used to consume the DC power.

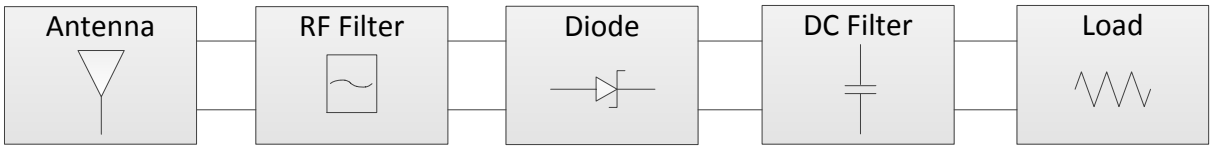


Figure 2.4: a block diagram for a generic rectenna system using an antenna as the primary collector.

The total efficiency of the rectenna system in most of the previously developed rectennas is obtained using the following relation [62]:

$$\eta_r = \frac{P_{out}}{P_{in}} \tag{2.2}$$

where the output power P_{out} and the input power P_{in} are given by:

$$P_{out} = \frac{{V_L}^2}{R_L} \tag{2.3}$$

$$P_{in} = \underbrace{G_r \left(\frac{\lambda^2}{4\pi}\right)}_{\text{Effective Area}} \times \underbrace{\left(\frac{G_t P_t}{4\pi D^2}\right)}_{\text{Incident Power Density}} \tag{2.4}$$

where V_L and R_L are the voltage across and the resistance of the connected load. In Eq. 2.4, G_t is the transmitter gain, P_t is the transmitted power and G_r is the gain of the receiver antenna. In measurements, the rectenna under test is usually placed a distance D away from the transmitter to ensure the far field condition is satisfied.

It is of interest to note that Eq. 2.2 represents the AC-to-DC conversion efficiency, which mainly depends on the type of diode used, the input power level and the optimal

resistive load. Hence, such equations do not take into account the efficiency of the collector (*i.e.,* antenna) where the microwave energy is converted to AC power in the first place. Therefore, a new method for computing the efficiency of rectenna systems is introduced in Chapter 4. In this method, the efficiency of the collector (microwave to AC link) is considered which is missing from earlier work.

Chapter 3

Design and Fabrication of 2.45 GHz Rectenna System

3.1 Introduction

This chapter presents the procedures for designing a full rectenna system operating at 2.45 GHz. First, a patch antenna fed by a microstrip line is designed and fabricated. Then a rectifier circuit consisting of a quarter wavelength transformer, a schottky diode, and an output filter is designed for an input impedance of 50 Ω to maximize energy transfer when attached to the patch antenna. Here, the primary goal of designing a rectenna system is not maximizing system efficiency but rather understanding the design procedures and the challenges one can face when building a rectenna system.

3.2 Microstrip Antenna

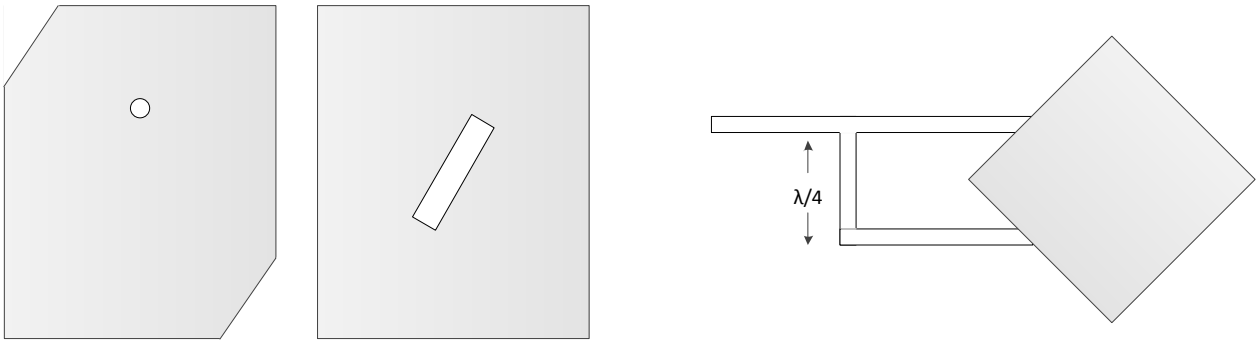
3.2.1 Theoretical Background

An antenna is one of the fundamental components in any rectenna system since it is the primary energy source that feeds the rectification circuit. Microstrip-type antennas are

widely used in communication systems due to their lightweight, low cost, ease of fabrication, low profile, conformability to planar and non-planer surfaces and mechanical robustness. However, microstrip antennas suffer from various drawbacks that limit their applications significantly such as narrow bandwidth, low power handling capability, high feed network losses and high cross polarization [23]. A great deal of research has been aimed at improving antenna characteristics and overcoming the aforementioned drawbacks of microstrip antennas. The advancements in recent antenna technology and the emergence of new concepts such as metamaterials have overcome most of these drawbacks or at least mitigated them to a certain extent. Therefore, antennas in the majority of rectenna systems if not all are of a microstrip type.

Microstrip antennas consist of a metallic layer, dielectric substrate, ground plane and feeding mechanism. Choosing the proper feeding configuration is important to achieve better impedance matching to the rectification circuit. The most common types, used in practise, are the coaxial probe, microstrip line, proximity coupling and aperture coupling [9]. Here, the microstrip line feeding mechanism is used since both the antenna and the rectification circuit can be easily fabricated in one plane. In addition, the microstrip line method is simple to match since the input impedance of the antenna can be varied by controlling the position and the width of the transmission line.

Recent rectenna research has focused on using circularly polarized (CP) microstrip antennas. Rectennas that utilize CP antennas have the advantage of constant output voltage, which is essential in applications where the connected load is rated at certain voltage levels. Furthermore, CP antennas can receive the incoming signal regardless of their position. This capability is particularly important in RFID tags where an interrogator can identify or track a certain tag irrespective of its orientation. A microstrip antenna by its nature is linearly polarized (LP) since it is designed to operate at a single mode. For an antenna to radiate circular polarized waves, two orthogonal modes must be excited, having equal magnitudes but with a 90° phase shift. This condition can be achieved by introducing some modifications to the antenna design. The common two methods to produce CP antennas are single and dual-orthogonal feeding mechanisms. CP can be obtained with a single feed by truncating the corners of a patch or by slightly perturbing some segments of the antenna as in Figure 3.1a. The main objective of these modifications is to disturb



(a) Single fed circular polarized antennas.

(b) Dual fed circular polarized antenna.

Figure 3.1: Single and dual fed circular polarized antennas.

the current density in such a way that the two modes are excited. In the dual-orthogonal feeding method, the two modes are excited by employing an external power divider circuit as in Figure 3.1b. To understand the type of polarization of any antenna, a plot of Axial Ratio (AR) vs. Frequency is used. An AR of less than 3 dB is indicative of an antenna radiating circular polarized waves within a certain range of bandwidth. The transmitter and the receiver antennas must have the same polarization, to avoid polarization mismatch and to maximize the energy collected by the receiving antenna.

3.2.2 Antenna Design

Since the goal of the work presented in this chapter is to clarify the working mechanism of rectenna systems, the basic LP microstrip-line fed patch antenna is selected.

A patch antenna is a resonance-based structure having a resonance frequency that depends on the various dimensions of the antenna, as shown in Figure 3.2. The width W and the length L can be found using the following relation [9]:

$$W = \frac{\upsilon_\circ}{2f_r}\sqrt{\frac{2}{\epsilon_r + 1}} \tag{3.1}$$

$$L = \frac{\lambda}{2} - 2\Delta L, \tag{3.2}$$

where $\upsilon_\circ$ is the speed of light, ϵ_r is the dielectric constant, and f_r is the resonance frequency of the antenna. The extended incremental length ΔL due to the fringing effect can be found from

$$\Delta L = 0.412(h)\frac{(\epsilon_{reff} + 0.3)\left(\frac{W}{h} + 0.264\right)}{(\epsilon_{reff} - 0.258)\left(\frac{W}{h} + 0.8\right)} \tag{3.3}$$

where h is the height of the substrate, and the effective dielectric constant ϵ_{reff} is given by the following:

$$\epsilon_{reff} = \frac{(\epsilon_r + 1)}{2} + \frac{(\epsilon_r - 1)}{2}\left[1 + 12\frac{h}{W}\right]^{-\frac{1}{2}}_{\text{for W/h >1}} \tag{3.4}$$

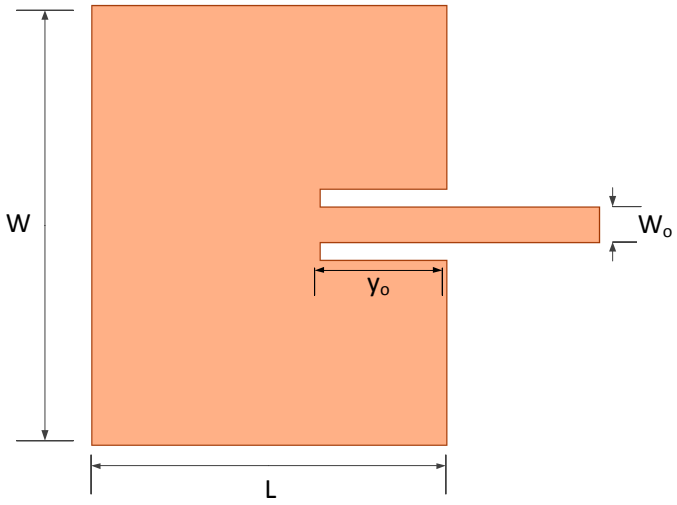


Figure 3.2: A schematic of the patch antenna showing all of its dimensions.

In Figure 3.2, the feed position $y_\circ$ and the width of the transmission line $W_\circ$ can be found using a model presented in [9]. Hence, the antenna parameters were found by solving for equations 3.1 - 3.4 and then using a full wave simulator HFSS to verify the results. Since the above set of equations will result in only an approximate result, one can fine tune the

Antenna Parameter	Value
L	31.8 mm
W	39.3 mm
$y_\circ$	11.3 mm
$W_\circ$	3.18 mm
ϵ_r	3.55
f_r	2.45 GHz

Table 3.1: A summary of the patch antenna parameters.

resonance frequency by slightly varying L and W to arrive at the desired frequency. Table 3.1 summarizes the final results of all the antenna parameters at an operating frequency of 2.45 GHz.

3.2.3 Antenna Fabrication and Measurement

The fabrication process is one of the vital stages learnt during the course of this work. The patch antenna and other structures used in this thesis were fabricated using a T-Tech milling machine available at the University of Waterloo. The fabricated patch antenna, illustrated in Figure 3.3, was etched on an RO4003 Rogers material having a relative permittivity of 3.55.

The return loss of the etched patch was measured using a Vector Network Analyzer (VNA) and, compared to the results obtained from HFSS, the measured return loss is in a good agreement with a slight frequency shift of 0.05 GHz, as shown in Figure 3.4.

3.3 Rectifier Circuit

This section discusses the main components of the rectifier circuit which will be attached to the fabricated antenna to form a full rectenna system. A number of rectifier topologies used in earlier work is reviewed. Furthermore, the input impedance of the agilent Schottky

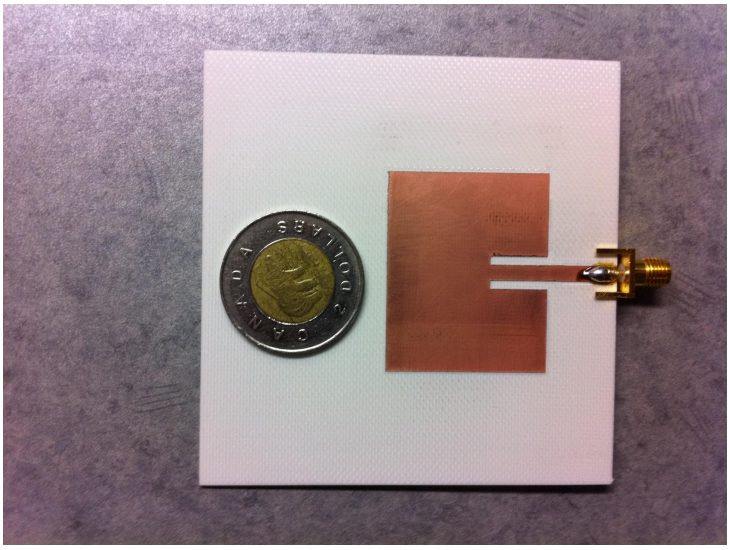


Figure 3.3: The fabricated microstrip-line fed patch antenna.

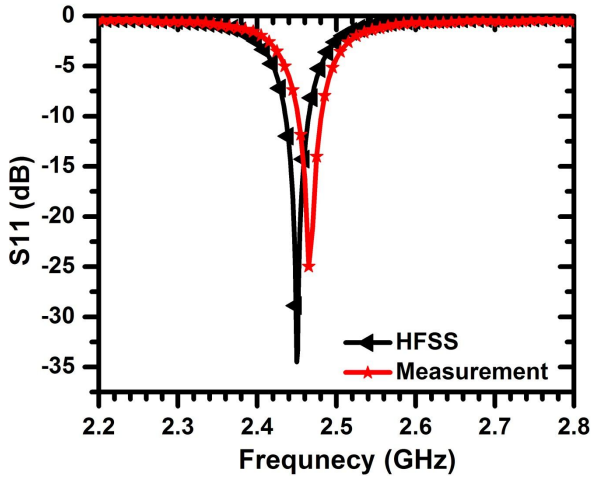


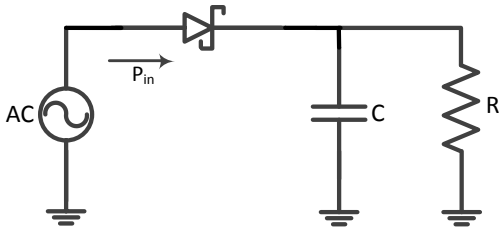
Figure 3.4: Return loss of the fabricated and simulated patch antenna.

diode used is determined using a diode model developed by K. Chang's group. Finally, the designs of the RF and DC filters are presented.

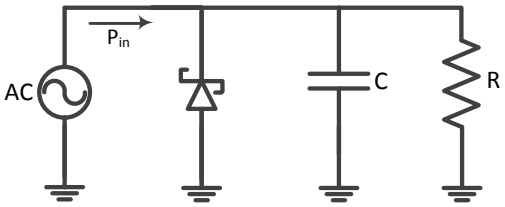
3.3.1 Rectifier Topologies

The main function of the rectifier circuit is to convert AC to DC power using a diode. Different from low frequency rectifiers, microwave rectifiers utilize surface mount diodes that operate at higher frequencies and matching networks to maximize power transfer before and after the diode. Depending on the targeted application, the rectifier circuit can employ a single or dual diodes. The three most common rectifier topologies used in the literature are the series diode, shunt diode, and Villard's voltage doubler rectifier circuits [58]. The series diode rectifier circuit consist of a diode connected in series, a capacitor and a load, as shown in Figure 3.5a. The diode blocks the negative cycle of the captured AC power, then the RC circuit smoothes out the rectified power to obtain a constant output voltage with low ripple. The series diode type of rectifier is ideal for rectennas that use a single feed antenna such as the microstrip-line fed and probe-fed patch antennas because of the simplicity of matching the diode's input impedance to that of the antenna by using a quarter wavelength transformer. The series diode rectifier is commonly used in rectenna applications such as energy harvesting [15] and RFID [55]. Similar to the series type, the shunt diode rectifier circuit consists of a capacitor, a load, and a diode, but with a parallel connection, as illustrated in Figure 3.5b. The shunt diode rectifier has the freedom of using a planar or co-planar microstrip line to link the various rectifier components. This type can usually be found in applications that utilize dual patch antennas, as in [41], or dipole antennas, as in [36]. The Villard's voltage doubler contains dual capacitors, dual diodes and a load, as shown in Figure 3.5c. As the name suggests, the Villard's voltage doubler achieves higher voltage as compared to the single diode type presented above. This can be obtained by using two cascaded diodes that rectify the AC power then, charge the dual capacitors. The Villard's voltage doubler rectifier-circuit was used in [44, 59] for energy scavenging applications.

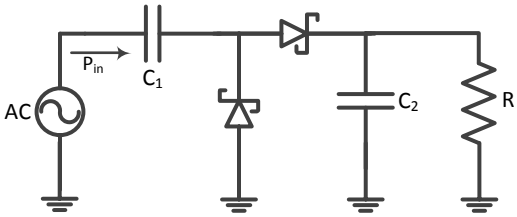
Here, the series diode rectifier circuit is used since it is the most suitable design for matching the microstrip-line fed patch antenna to the diode. In the following subsections, the type of diode used in the design and the matching networks are discussed.



(a) Series-diode rectifier circuit.



(b) Parallel-diode rectifier circuit.



(c) Villard's voltage doubler rectifier circuit.

Figure 3.5: Various rectifier-circuit topologies used in the literature [58].

3.3.2 Schottky Diode

When operating at higher frequencies (*i.e.,* 2.4 GHz), the common P-N junction diode is no longer capable of rectifying the AC signal since it has relatively slower switching capability and higher voltage drop. Instead, a Schottky diode is used to overcome these limitations. Unlike the P-N junction diode, which has a semiconductor-semiconductor junction, the Schottky Barrier Diode (SBD) is formed by plating a metal contact onto an n- or p-type semiconductor [17]. This metal-semiconductor junction acts as a diode, allowing the current to flow in one direction and blocking it in the other. The I-V curve of a Schottky diode is similar to that of a P-N junction diode, with the exception of [47]:

- Faster switching between conducting and non-conducting modes due to the absence of the minority-carrier charge storage effect
- Lower forward voltage drop

The two main classes of Schottky diodes are the n-type and the p-type Schottky diodes. The n-type Schottky diode is formed by a metal and type n semiconductor junction. This type is distinguished by having a high barrier height (indicating low saturation current) and a low series resistance. The n-type Schottky diode is used in applications where a DC bias is accessible, such as mixer and detector applications. The p-type, on the other hand, is made up of a metal and type p semiconductor, and it is characterized by a relatively low barrier height and high series resistance. This type of Schottky diode is used in applications where DC bias is not available, such as square law detector applications [54]. The diode model and proper diode-type selection are presented in the next section.

3.3.3 Diode Model

One of the most important initial steps when designing a rectenna system is determining the input impedance of the Schottky diode. This is due to the non-linear behaviour of the diode as its input impedance depends significantly on the input power level, the load impedance and the frequency of operation. Here, the diode model developed by K. Chang's

group is used due to the promising results and the great agreements between the measured and the simulated results achieved in the literature, as in [50]. For simplicity this model does not take into account the loss due to higher-order harmonics.

The equivalent circuit of a Schottky diode contains a series resistance R_s, a junction resistance R_j, and a junction capacitance C_j as presented in [34]. The load resistance R_L can be varied along with the input power level to arrive at the desired diode input impedance.

The input impedance of the diode Z_d can be found by applying Kirchoff's voltage law to the equivalent circuit of the schottky diode. The final form of the Z_d is given by [34]:

$$Z_d = \frac{\pi R_s}{\cos\theta_{on}\left(\frac{\theta_{on}}{\cos\theta_{on}} - \sin\theta_{on}\right) + j\omega R_s C_j\left(\frac{\pi-\theta_{on}}{\cos\theta_{on}} + \sin\theta_{on}\right)} \tag{3.5}$$

The turn on angle θ_{on} depends on the input power level and can be found from the following relation:

$$\tan\theta_{on} - \theta_{on} = \frac{\pi R_s}{R_L\left(1 + \frac{V_{bi}}{V_\circ}\right)} \tag{3.6}$$

where $V_\circ$ is the output voltage across the load and V_{bi} is the built-in voltage. The junction capacitance C_j is given by:

$$C_j = C_{j\circ}\sqrt{\frac{V_{bi}}{V_{bi} + |V_\circ|}} \tag{3.7}$$

A detailed derivation of Eq. 3.5 can be found in [34].

The Schottky diode selected for the rectenna under test is the Avago's HSMS2860 Schottky diode. This type of diode is optimized to operate at a frequency range of 915 MHz-5.8GHz. The diode has a breakdown voltage V_{br} of 7.0 V and a forward voltage of 0.65 V. The zero bias junction capacitance $C_{j\circ}$ and the series resistance R_s are 0.18 pF and 6.0 Ω, respectively.

A Matlab code, which makes use of the integrated optimization toolbox, is developed to solve for Z_d, θ_{on}, and C_j. Figure 3.6 shows the real part of Z_d, as the load resistance is varied from 1-1000 Ω. Z_d was found to be 170.8 Ω when the load resistance is 250 Ω. Here the choice of the load resistance is kept low to ease the fabrication process as a higher load resistance will lead to a narrower microstrip-line and ,hence, harder to fabricate. The output voltage is usually taken to be 50% of the breakdown voltage to prevent burning the diode, which would create an open circuit. Therefore, the output voltage $V_\circ$ is set at 3.5 V.

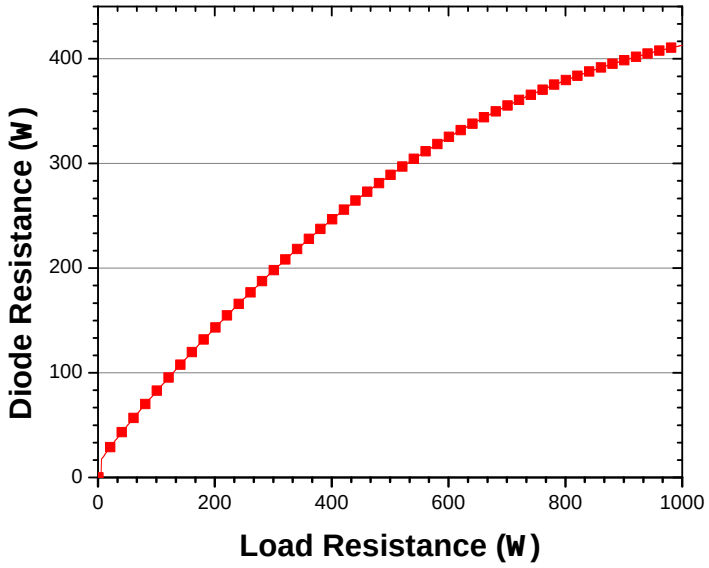


Figure 3.6: Load resistance Vs. Diode input resistance.

3.3.4 RF and DC Filter

At this stage, the two main components of the rectenna system have been designed and analysed and need to be connected through a matching network since they have different input impedances. The primary functions of the pre-rectification or RF filter are to match

the real part of the diode's input impedance to that of the antenna and to prevent higher-order harmonics caused by the non-linear behavior of the diode from radiating by the antenna. A quarter wavelength transformer is inserted between the antenna and the diode to transform the diode resistance to 50 Ω. The impedance of the transformer can be found from the following relation:

$$Z_{QWT} = \sqrt{Z_A Z_d} \tag{3.8}$$

where Z_A is the input impedance of the antenna. Using the diode resistance obtained from Figure 3.6, Z_{QWT} is found to be 92.41 Ω. Hence, using a microstrip-line calculator, the length and width of the quarter wavelength transformer are $l = 16.7\ mm$ and $w = 1.05\ mm$.

Finally, the DC or post-rectification filter is used to tune the reactance of the diode impedance and to prevent any higher frequencies from reaching the load. The DC filter generally consists of a series microstrip line and a shunt capacitor. The Advance Design System (ADS) simulator is used to optimize the capacitor value, and the length and width of the transmission line such that the magnitude of the reflection coefficient (S11) looking from the 50 Ω port into the rectifier circuit is minimized (*i.e.,* a magnitude of $S11 \approx 0$).

From the simulation, it is found that a capacitor value of $C = 56\ pF$ and a transmission line with $l = 40\ mm$ and $w = 1.9\ mm$ would yield maximum matching between the diode and the load.

Figure 3.7 shows the fabricated rectifier circuit with an LED that mimics a load. The rectifier was tested with a high frequency power source; however, the LED did not glow since not enough voltage was developed across it to turn it on.

3.4 Conclusion

The procedures for designing a rectenna system from the antenna to the load were presented. The input impedance of the Agilent diode was determined from a Schottky diode model available in the literature. A quarter wavelength transformer was designed to match

the real part of the diode to that of the antenna. Then, using an ADS simulator, a DC filter was designed to tune out the reactance of the diode impedance. A surface mount LED was used instead of the load as an indicator to test the rectifier circuit. However, the LED did not turn on as the rectifier circuit was not able to deliver enough voltage to the LED. To overcome this issue, some modifications can be made, such as the use of a Villard's voltage doubler circuit for higher voltage levels, and the use of better collector to capture more energy, as will be discussed in the following chapter. Even though the LED did not turn on, the goal for this chapter was achieved since this work is meant to be a preliminary for future work explained in the next chapter.

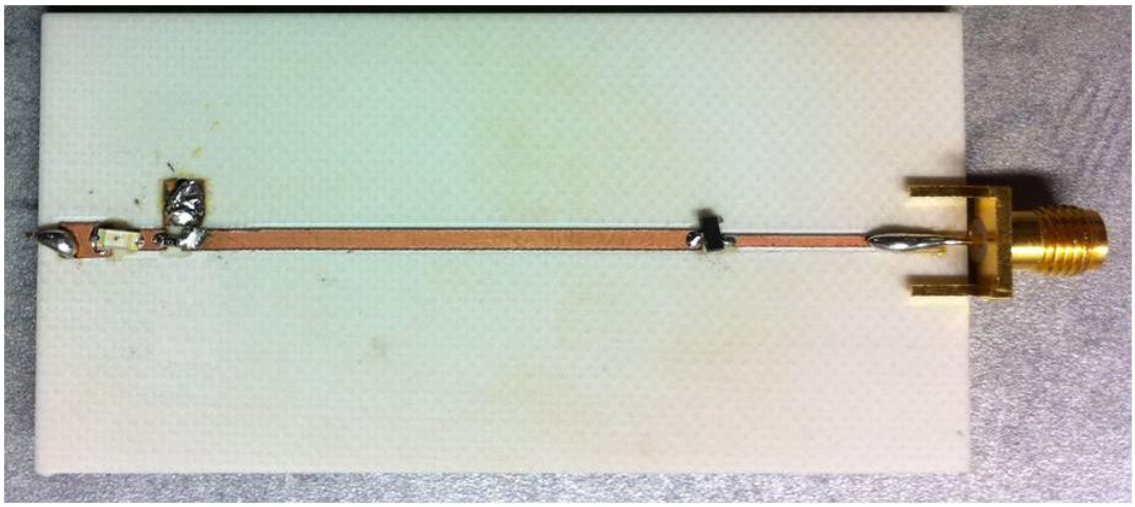


Figure 3.7: Fabricated rectification circuit.

Chapter 4

Electromagnetic Energy Harvesting Using Metamaterial Particles

4.1 Introduction

This chapter presents a novel energy collector based on metamaterial particles in what is known as a Split Ring Resonator (SRR). The subject of metamaterials is briefly introduced, then a feasibility study of SRRs to harvest energy is presented. To prove the concept, a 5.8 GHz SRR is designed and fabricated and then tested using a power source, an Infiniium oscilloscope and a commercially available patch antenna array. In addition, a new efficiency concept is introduced, taking into account the microwave-to-AC conversion efficiency which is missing from earlier work. Finally, a 9X9 SRR array is compared with a 2X2 patch antenna array, both placed in a fixed footprint.

4.2 Metamaterial Background

The emergence of metamaterials has drawn considerable attention from physicists and engineers researching in a wide range of streams. The fact that metamaterials can produce media with negative permittivity and/or permeability has originated numerous or

enhanced existing applications such as creating perfect lenses and absorbers [31, 39], decoupling and modifying antenna characteristics [8, 20, 5], improving the efficiency of solar cells, and detecting cracks [4], to name but a few. Metamaterials can be defined as "artificial media with unusual electromagnetic properties" [49]. Metamaterial particles are formed by assembling electrically small resonators that can take various shapes, geometries and compositions. One of the most common types of resonators is the SRR, which is a metallic inclusion with one or more splits. An SRR can be made of (single or multiple) and (concentric or parallel) electrically small rings. It also can take various shapes, such as those shown in Figure 4.2, or many others analysed in the literature. This work investigates the feasibility of only a single loop resonator to harvest energy, but generalization can be made to any structure available in the literature. When modelling a single-loop SRR (Figure 4.1), it can be realized as a simple LC circuit where the size of the gap (gxw) and the arm length of the metallic ring (L) contribute mainly to the total capacitance and inductance of the structure, respectively. Hence, by varying such dimensions, one can design an SRR to resonate at the desired frequency. A detailed study of the effect of varying the mentioned parameters on the SRR's resonance frequency can be found in [6]. The next section presents a study of the feasibility of such a resonator to harvest electromagnetic energy.

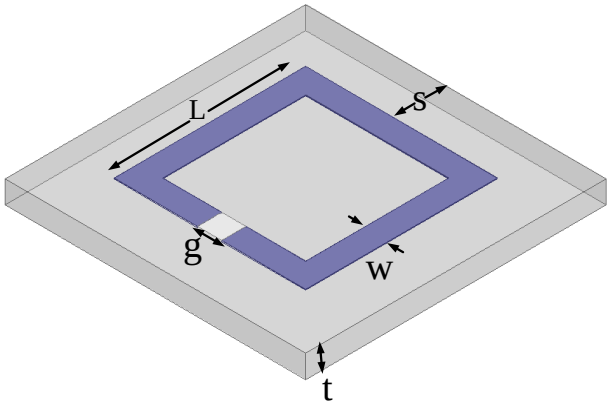


Figure 4.1: Split Ring Resonator unit cell.

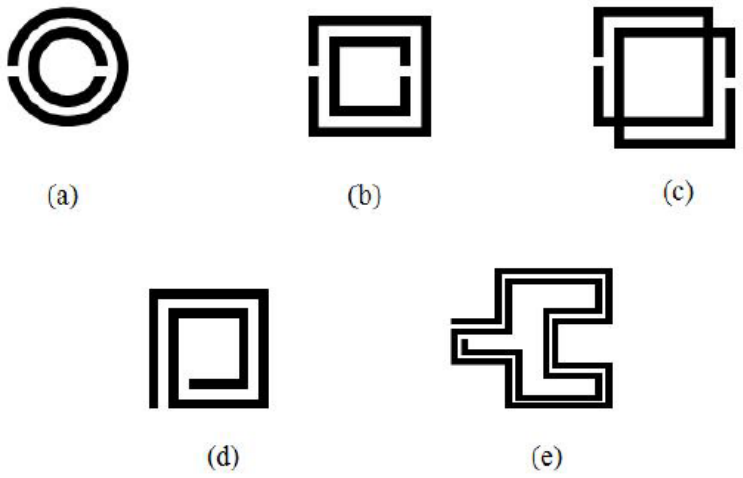


Figure 4.2: Various shapes of metamaterial particles proposed in the literature [7].

4.3 Feasibility of SRRs to Harvest Electromagnetic Energy

When an SRR is excited by a magnetic field normal to its plane, a highly concentrated electric field is developed across the gap of the structure, as indicated by the red color in Figure 4.3. Equally interesting is that the SRR can be excited by an incident magnetic field with various polarizations, as discussed in section 4.5. This concentration of a high electric field is indicative of a build up of high voltage across the gap of the SRR. Hence, this work proposes harvesting this energy by means of a load placed across the gap. To realize this concept using HFSS, a resistive sheet is placed across the gap of an SRR cell, resonating at 5.8 GHz. The designed SRR has dimensions of l =5.9 mm, w = 0.55 mm and g= 0.8 mm. Since the optimal resistance value is not known, the resistive sheet is assigned a variable resistance value ranging between 100 Ω - 10,000 Ω. The SRR is excited by a plane wave such that the H field is perpendicular to the SRR plane. The efficiency of the SRR is then calculated by using the proposed efficiency concept presented in section 4.5. It was found that a single SRR cell has an efficiency of around 40%, with an optimal

resistive load of 2.3 KΩ. This result suggests that the voltage developed across the gap is mostly dissipated by the resistive sheet. Therefore, such SRR structures can potentially be used for harvesting electromagnetic energy.

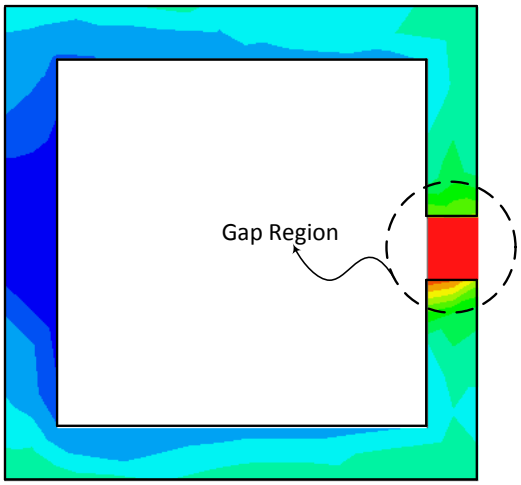
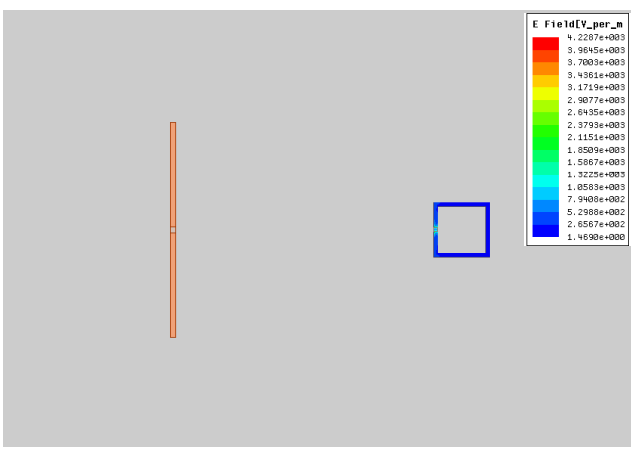


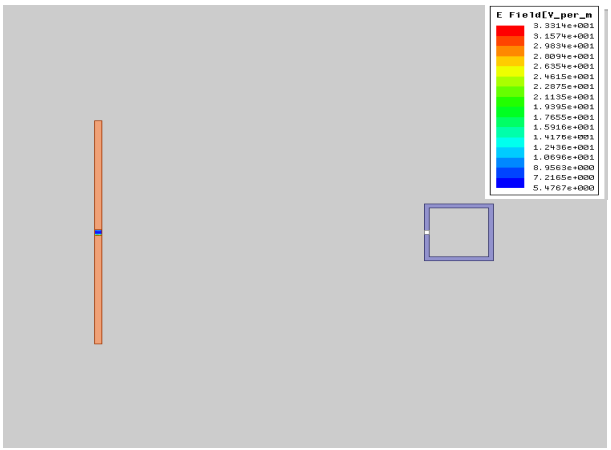
Figure 4.3: Electric field distribution on the plane of the SRR at resonance frequency.

For any radiator to receive energy, it must obey the reciprocity theorem. This theorem states that "in any network composed of linear, bilateral, lumped elements, if one places a constant **current***(voltage)* generator between two **nodes** *(in any branch)* and places a **voltage** *(current)* meter between any other two **nodes** *(in any branch)*, makes observation of the meter reading, then interchanges the locations of the source and the meter, the meter reading will be unchanged" [9]. To ensure that the theorem is not violated, an experiment in HFSS is conducted by designing two radiators, a dipole antenna and a single loop SRR both resonating at the same frequency. The experiment is divided into two cases:

1. A dipole antenna is excited by a current source placed at its feed; then the voltage across the gap of the SRR is recorded (Figure 4.4a).



(a) A snapshot of the simulation for case 1.



(b) A snapshot of the simulation for case 2.

Figure 4.4: A simulation that verifies the reciprocity therm using a dipole antenna and an SRR.

2. An SRR is excited by a current source placed across its gap; then the voltage across the feed of the dipole antenna is recorded (Figure 4.4b).

The voltage of both cases can be found by

$$V = E \times d \tag{4.1}$$

where E denotes the electric field, and d is the length of the feed (for the dipole) and the length of the gap (for the SRR). Therefore, the voltages for both cases are

$$V_1 = E_1 \times d_1 = (5.988 \times 10^4) \times (0.8 \times 10^3) = \mathbf{47.907\ V}\ \text{for case 1} \tag{4.2}$$

$$V_2 = E_2 \times d_2 = (3.8562 \times 10^4) \times (1.23 \times 10^3) = \mathbf{47.43\ V}\ \text{for case 2} \tag{4.3}$$

It is evident from the voltage values of both cases that the SRR obeys the reciprocity theorem and therefore can be used for collecting electromagnetic energy.

4.4 Proof of Concept

Now, that the simulation analysis has resulted in promising conclusions, the feasibility of using an SRR to harvest electromagnetic energy is validated by testing and measurements. First, the single loop SRR simulated above was fabricated using a Rogers Duroid RT5880 substrate with thickness of 0.79 mm. Then the SRR was loaded with a surface mount resistor of 2.7 KΩ as shown in Fiqure 4.5. Here, the resistor used in the experiment is different from that of the optimal resistor (2.3 KΩ) obtained from the simulation since the latter was not available at the time of the experiment. An experiment was then conducted using the following measurement setup (Figure 4.6):

1. A commercially available 19 dBi gain array antenna operating at 5.8 GHz.
2. An Agilent Infiniium 91304ADSA 12 GHz oscilloscope equipped with a single-ended probe.
3. A high frequency 30 dbm power source.
4. The fabricated single loop SRR.

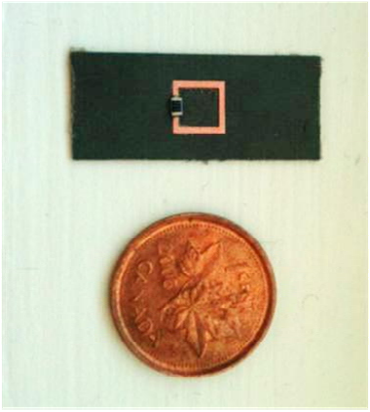


Figure 4.5: Fabricated SRR loaded with a 2.7 KΩ resistor placed at the gap.

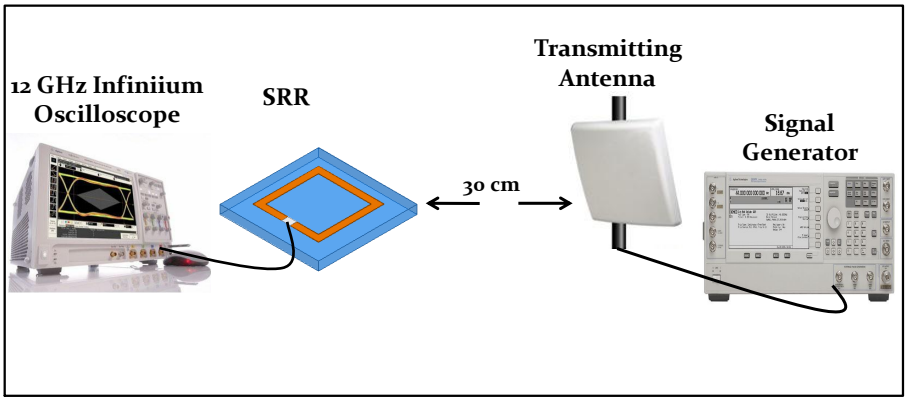


Figure 4.6: Experimental setup.

The SRR was placed 30 cm away from the antenna, and was positioned in such a way that the H-field of the illuminated wave was perpendicular to the plane of the structure. The antenna was excited by a power source with a power level of 24 dbm. Then the voltage across the resistor of the SRR was measured using a single-ended probe of the oscilloscope. A snapshot of the voltage readings (Figure 4.7) shows that the voltage measured across the resistor was 611 mV.

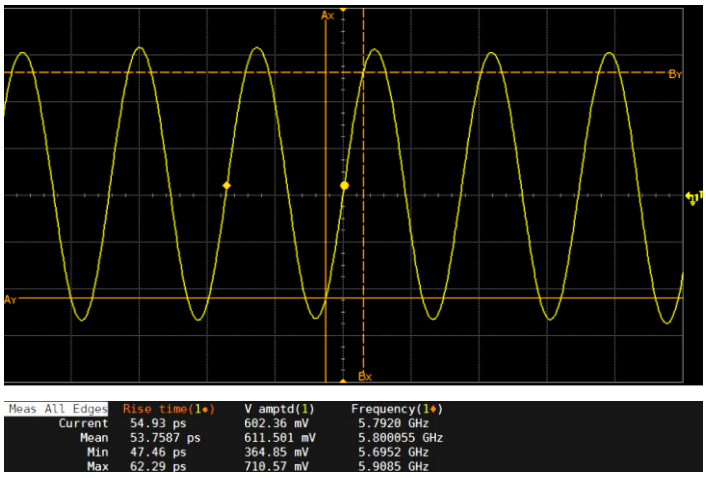


Figure 4.7: A snapshot of the voltage across the gap of the SRR.

4.5 SRR Array Vs. Patch Antenna Array

In this section, a new method for calculating the efficiency of electromagnetic energy collectors such as antennas and SRRs is proposed. Then a comparison in terms of the total microwave-to-AC conversion efficiency between an array of SRRs and an array of antennas placed in the same footprint or area is presented.

As mentioned in Chapter 2, the efficiency of rectenna systems reported in literature does not take into account the microwave-to-AC conversion efficiency of the energy collector. Here, what is meant by microwave-to-AC link is The efficiency of the antenna to convert all the power incident on a specific footprint to available power at its feed position. Therefore, a footprint in square meters must be defined, over which a number of collectors is placed in such a way that the power collected is maximum. Hence the efficiency of a collector as defined above can be found as follows:

$$\eta = \frac{P_{av}}{P_{area}} \tag{4.4}$$

where P_{area} is the total time-average power incident on the footprint, and P_{av} is the the maximum available time-average ac power received by the collector or all collectors occupying the specific footprint under consideration and is available at the feed terminal of the receiving collector. Therefore, P_{av} is given by the following relation:

$$P_{av} = \sum_{i=1}^{n} \frac{V_i^2}{R_i} \tag{4.5}$$

where V_i and R_i are the voltage across and the resistance of collector i. The total number of collectors on a specific footprint is denoted by n.

Considering the efficiency definition just introduced, a demonstration is presented comparing the efficiency of an array of SRRs with an array of patch antennas placed in the same footprint as shown in Figure 4.9. The array of SRRs contained 81 single loops, all of identical size and designed to resonate at around 5.85 GHz. In addition, an array of 2 X 2 identical patch antennas was placed in the same footprint, each resonating at the

same frequency of around 5.85 GHz. The total footprint area was 85 X 85 mm^2. It was reported in the literature that antennas need to be separated by at least $\lambda/2$ to retain their characteristics such as radiation pattern and gain [32]. Therefore, the 4 antennas were separated by a distance of around $\lambda/2$ to ensure maximum power collection by the antennas. In addition, each antenna was fed by a coax probe from beneath. The performance of a probe-fed patch antenna is greatly dependent on the feed position. Hence, the feed position was first analysed by varying the location of the coax with a distance r away from the center of the patch and along the axis that is parallel to the largest dimension of the patch antenna, as shown in Figure 4.9. It was found that the best performance for all of the 4 antennas was achieved when the probe was placed a distance of $r = 2$ mm away from the center of the antenna, as shown in Figure 4.8.

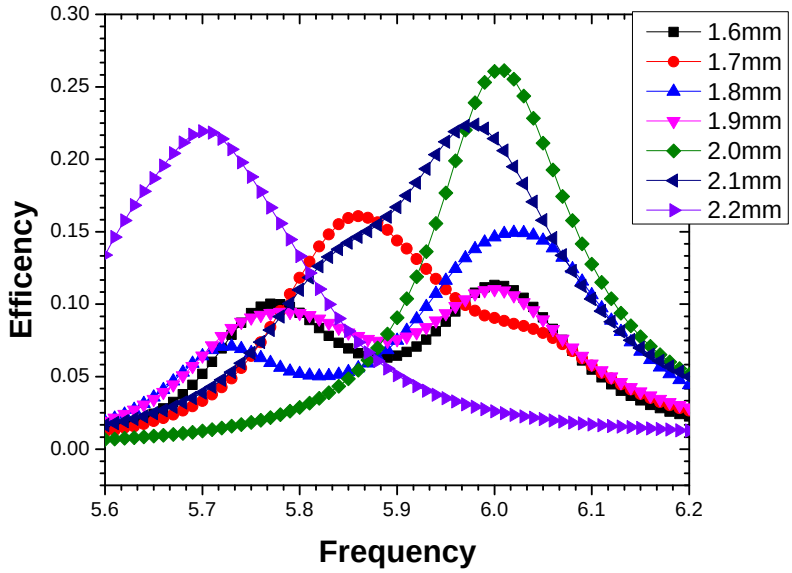


Figure 4.8: The efficiency of the patch antenna with varies coax-probe positions.

In the demonstration, each array was excited by a horn antenna placed a distance of 120 cm away from the array to ensure that the far field condition was satisfied, and a plane wave was incident on the array. Since both the antenna and the SRR are polarized

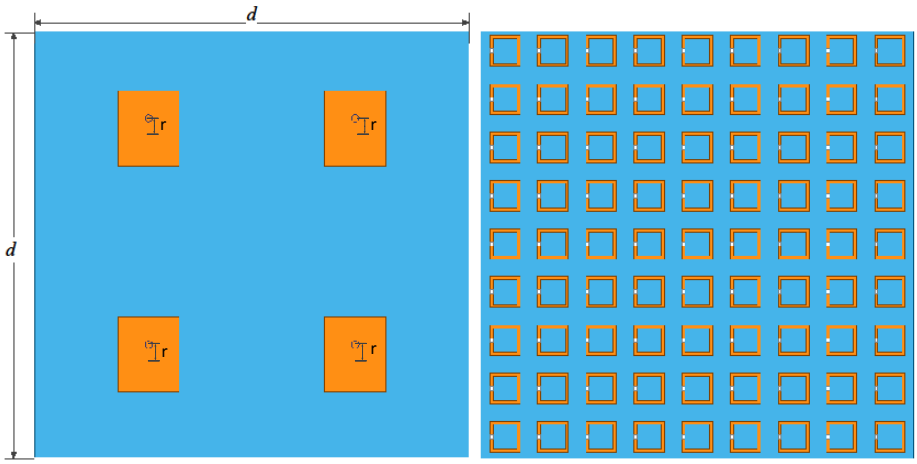


Figure 4.9: a 9 x 9 SRR array and a 2 X 2 patch antenna array occupying the same footprint.

differently, each array was tilted an angle ϕ with respect to an axis normal to the plane of the array, as indicated in Figure 4.10. Three tests were conducted for each array, with angles of 30°, 45°, and 60°. Figures 4.11, 4.12, 4.13 show the efficiency of the antenna array and the SRR array at an angle of $\phi = 30°$, 45° and 60°, respectively.

The following conclusion can be drawn:

- The SRR array resulted in higher efficiency in all three cases.
- The SRR structure is much smaller in size than the antenna in the specific footprint mentioned above, which can contain either 81 SRRs or only 4 patch antennas.
- The bandwidth over which the energy is collected is much wider than that of the antenna array.

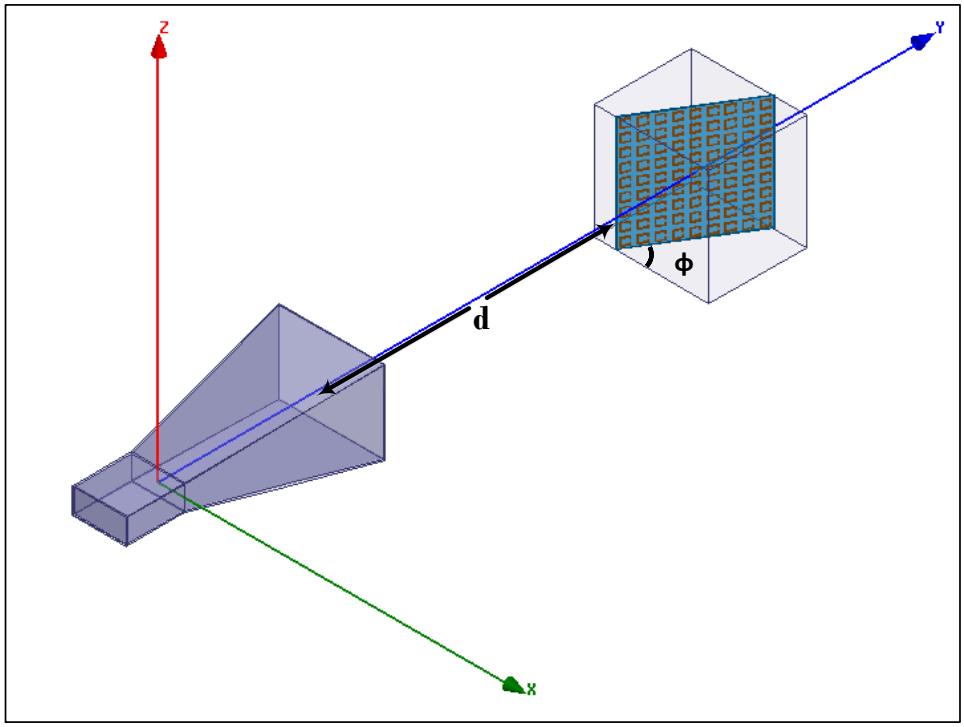


Figure 4.10: Simulation setup for energy harvesting using a horn antenna as the source of radiation and an SRR array as the collector.

4.6 Conclusion

This chapter proposed a new method using SRR cells for collecting electromagnetic energy. The feasibility of using SRRs to harvest microwave energy was studied through simulation by first placing a resistive load across the gap of the single loop SRR, then calculating the power dissipated across the resistor. It was found that around 40 % of the power incident on the single cell SRR is harvested and dissipated by the resistive load. In addition, a simulation experiment was conducted to ensure that the reciprocity theorem is not violated, and the SRR can indeed be regarded as a receiver. An experiment was then conducted to test the feasibility of SRRs to harvest microwave energy. The fabricated SRR cell was excited by an antenna placed 30 cm away, and then the voltage across the gap was measured using a high-frequency oscilloscope. A voltage of 611 mV was observed across the gap when

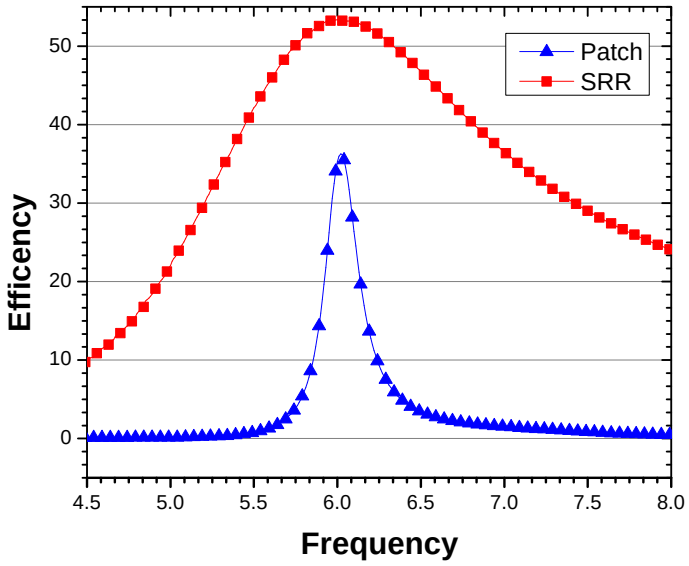


Figure 4.11: Energy harvesting efficiency of the 9 x 9 SRR array Vs. a 2 X 2 patch antenna array both tilted at 30°.

24 dbm power level was pumped into the antenna feed. Finally, the efficiency of an array of SRRs was compared to that of an array of antennas occupying the same footprint. The SRR array resulted in at least 25% more efficiency and a wider energy collection bandwidth than the antenna array.

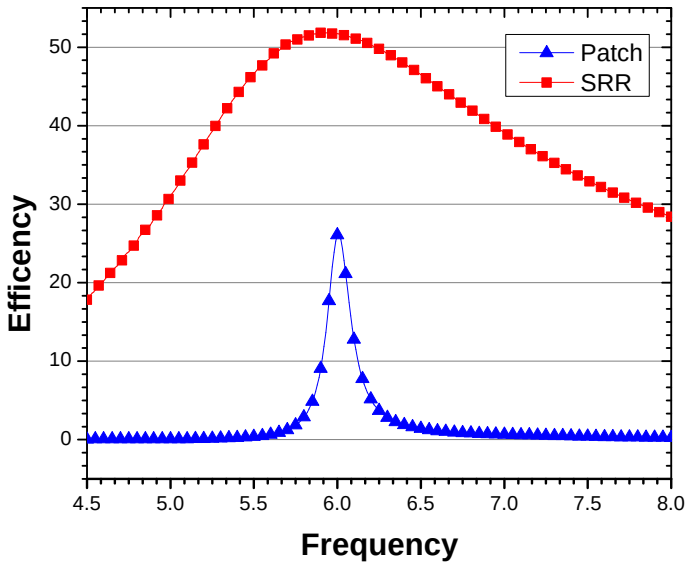


Figure 4.12: Energy harvesting efficiency of the 9 x 9 SRR array Vs. a 2 X 2 patch antenna array both tilted at 45°.

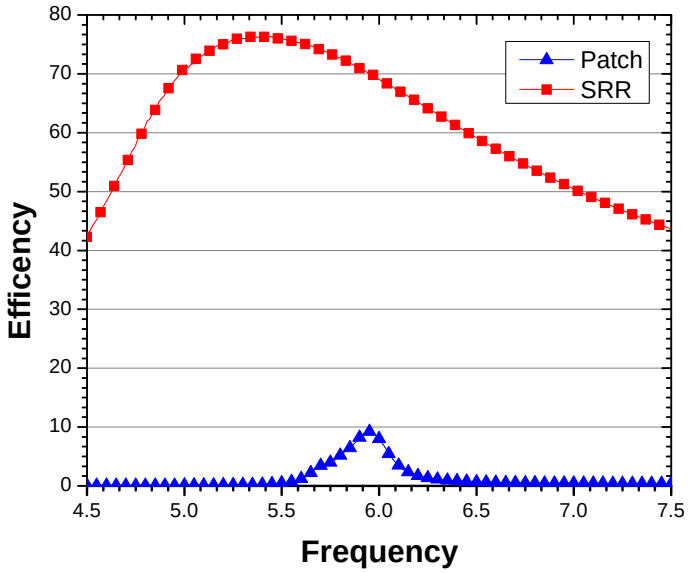


Figure 4.13: Energy harvesting efficiency of the 9 x 9 SRR array Vs. a 2 X 2 patch antenna array both tilted at 60°.

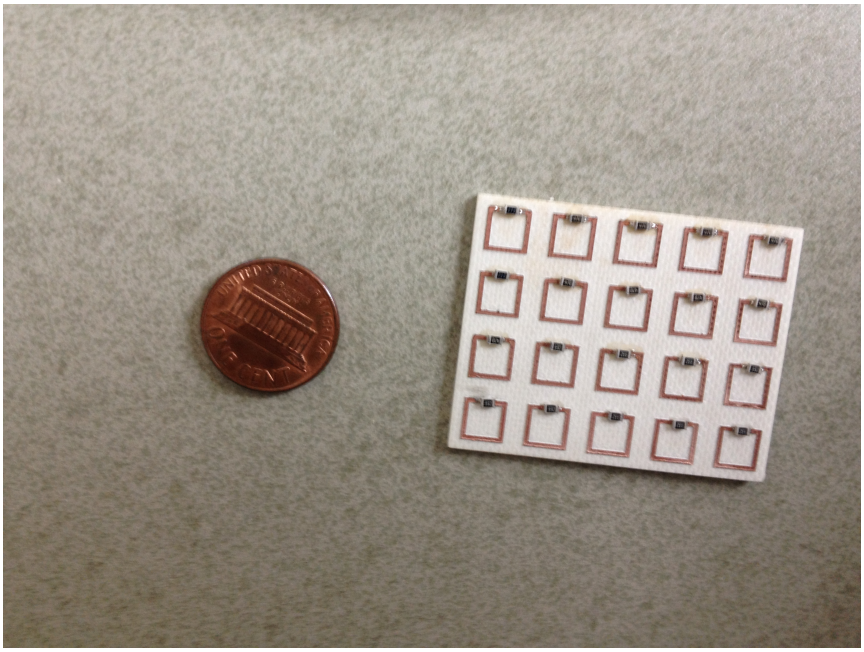


Figure 4.14: A sample of a fabricated 4 X 5 SRR array.

Chapter 5

Conclusion and Future Work

5.1 Conclusion

In this thesis, a review of energy harvesting systems, which in the literature lies under the topic of WPT and rectenna systems, was presented. Then, detailed procedures for designing and fabricating a full rectenna system were provided. A new energy collector based on metamaterial particles was proposed for the purpose of replacing the existing collector (antenna) that has lower efficiency and much larger size than an SRR unit cell. An experiment was conducted both in simulation and in measurements to examine the feasibility of SRRs to harvest microwave energy. The results from both tests show that such metamaterial structure can indeed be used as an energy harvester. Finally, the efficiency of an array of 9 X 9 SRRs was compared to that of an array of 2 X 2 patch antennas placed over an identical area. At least 25% more efficiency and a wider energy collection bandwidth were achieved by the SRR array.

5.2 Future Work

The following suggestions are worth pursuing for future work:

- Improve the design of the rectenna system to provide more output power and hence higher efficiency.
- Investigate the efficiency of various SRR structures such as the those shown in Figure 4.2.
- Study the effect of coupling on the efficiency of both antennas and SRRs.
- Integrate an SRR array, as the one shown in Figure 4.14, to a rectifier circuit for a full Microwave to DC conversion system.

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RECTENNA TECHNOLOGY PROGRAM: Ultra Light 2.45 GHz Rectenna and 20 GHz Rectenna

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16. Abstract

The program had two general objectives. The first objective was to develop the two plane rectenna format for space application at 2.45 GHz. The resultant foreplane was a thin-film, etched-circuit format fabricated from a laminate composed of 2 mil Kapton F sandwiched between sheets of 1 oz. copper. The thin-film foreplane contains half wave dipoles, filter circuits, rectifying Schottky diode, and DC bussing leads. It weighs 160 grams per square meter. Efficiency and DC power output density were measured at 85% and 1 kw/m^2, respectively. Special testing techniques to measure temperature of circuit and diode without perturbing microwave operation using the fluoroptic thermometer were developed.

A second objective was to investigate rectenna technology for use at 20 GHz and higher frequencies. Several fabrication formats including the thin-film scaled from 2.45 GHz, ceramic substrate and silk-screening, and monolithic were investigated, with the conclusion that the monolithic approach was the best. A preliminary design of the monolithic rectenna structure and the integrated Schottky diode were made.

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SUMMARY

This contract had two major objectives. The first was to refine the rudimentary technology of the 2.45 GHz thin-film, etched-circuit rectenna with particular emphasis upon its space applications. The second was to examine the kinds of rectenna technology best suited for rectenna operation at frequencies of 20 GHz and higher.

The status of thin film rectenna technology at the start of the study was a single individual rectenna element made from a laminate of mylar and one ounce copper. The rectenna element was inefficient and otherwise unsatisfactory. The current study revealed why the structure was inefficient and found that a laminate of Kapton F and copper was a much better material. The proper masks were designed and individual rectenna elements fabricated. The rectifying diodes were added and the individual elements tested in a closed system where an overall efficiency of 85% was achieved.

Then arrays of up to 30 rectenna elements on one continuous laminate were constructed. These arrays were thoroughly checked for power handling capability up to 5 watts average output per rectenna element. In addition, the power handling capability of individual elements was evaluated as a function of velocity of air flow over the rectenna surface. The diode temperature was simultaneously monitored by a unique, non-invasive test instrument, the fluoroptic thermometer. From this data, reliable estimates could be made of the power handling capability of a complete rectenna array as a function of air density and air flow velocity. A 25 element section was sent to LeRC for test and evaluation.

A preliminary study was made of the deployment of the rolled up rectenna into a flat plane for space use. Various problems were investigated and several formats examined.

The study of the various approaches to fabricating a high frequency rectenna revealed that the thin-film, etched-circuit rectenna format was not a sound approach and stimulated an investigation of placing the foreplane of the rectenna on a solid dielectric substrate such as alumina ceramic. This ceramic could serve as both a separator from and a heat conducting path to the reflecting plane where heat could be removed by flow of a coolant. Further study indicated that a monolithic structure based on the use of a GaAs substrate on which both diodes and circuits were formed would be the best approach for constructing a rectenna at frequencies of 20 GHz and higher. A diode was designed for the monolithic structure, and theoretically evaluated in terms of efficiency and power handling capability. An initial design of a monolithic rectenna was carried out.

1.0 INTRODUCTION

1.1 General Objectives

The contract had two general objectives. The original objective dealt exclusively with the development of the thin-film, printed-circuit rectenna for space use at a frequency of 2.45 GHz. A principal part of this objective was concerned with improving the basic properties of a rectenna format that had resulted from earlier embryonic efforts to develop a thin-film rectenna. The second general objective, which was the subject of a contract extension, was to examine the application of the rectenna principle to high frequency rectennas at 20 GHz and above.

For purposes of discussion it will be desirable to handle these two objectives separately. The first objective, that having to do with the improvement of the 2.45 GHz rectenna and its application to space will be discussed first.

1.2 Improvement of the 2.45 MHz Thin-Film, Etched-Circuit Rectenna and Its Application to Space

The more detailed objectives of the work carried out under this subject are described in the first four items in the statement of work given in Section 2.0. In the introduction we will discuss the work in more general terms.

The approach to carrying out the task was based upon an extension of an approach introduced under NASA contract NAS6-3006(1). This approach was based upon the use of conventional thin-film printed circuit technology, in which all elements of the circuitry, excluding the rectifying diode and the reflecting plane, were "printed" on the two surfaces of the film without the ncessity of any interconnects between the circuitry on the two surfaces. This fabrication method is consistent with the objective of producing very large areas of the rectenna at a low cost per unit area and with a high ratio of power output to mass of the structure.

However, the technology for the thin-film rectenna had proceeded only to the point of making and testing individual elements made from a laminate composed of mylar covered with thin sheets of copper. The efficiency was much less than expected, and the mylar was very vulnerable to both ultra violet degradation and high temperature operation.

The general work effort related to this first general objective consisted of analyzing and refining the thin-film, etched-circuit rectenna at the single rectenna element level for use in space and then fabricating relatively large areas of the rectenna for evaluation and test. Another aspect of the work related to storage of the rectenna while enroute to space and the deployment of it in space. Still another aspect was the aid given to LeRC in the design of a facility for testing the rectenna.

The results of the effort to improve the rectenna in a format more suitable for space were very successful. It was found that the use of Kapton F material as the core of the laminate material greatly extended the permissible operating temperature as well as making the rectenna highly resistant to deterioration from ultra violet ration. In addition it was determined that Kapton F greatly reduced the losses that were inherent in the mylar itself as well as in the lossy adhesive used to join the mylar to the copper.

One of the interesting developments related to the diode rectifier. No portion of the rectenna is more important than the diode. Fortunately, the basic development of the diode had already taken place during earlier rectenna work. The rectifier is a Schottky barrier diode that utilizes GaAs as the semiconductor material. The series resistive loss in this material is much less than that of silicon, the only other candidate material, so it is considerably more efficient. The diode also uses a heat sink that is fabricated on the metallic side of the Schottky barrier to provide a low impedance path for heat flow from the active and heat generating portion of the diode.

The packaging of the diode has taken several formats. The first successfully used package was of the ceramic pill type. In the interests of greatly reducing the production cost of the diode for use in the Solar Power Satellite application a glass packaging technique was introduced. It had been expected to use this packaging format for the thin-film rectenna; in fact, it was used in the early phase of the work effort. However, the tooling for the glass diode became unavailable and it was necessary to shift back to a ceramic package. On subsequent analysis it was discovered that the thermal conduction from such a package is about twice that from the glass diode so that heat is conducted more efficiently to the printed circuit which presents the surface from which the heat is radiated to space. The result is that the power handling capability of a rectenna built from such diodes is considerably better than anticipated, resulting in the prospect of the rectenna working at considerable higher power density in space than originally anticipated.

It was not feasible within the constraints of the contract funding to test the rectenna in vacuum. However, the opportunity did arise to check the rectenna under known rates of a convective flow of air, while simultaneously monitoring the temperature of the diode. The results of this investigation were important in feasibility studies of microwave powered aircraft that would fly at high altitudes where the air was much less dense but where the aircraft flight speeds resulted in convection cooling rates similar to those taken in the laboratory under sea level air conditions. Furthermore, one data point in the studies was for zero convective air flow. Even under these conditions the temperature rise in the diode remained below 100°C for two watts of DC power output per rectenna element and diode.

It was possible to make the interesting results of this experimental evaluation of the power handling capability and efficiency of the thin film rectenna publicly available in a timely fashion through two oral presentations

and printed papers at two international microwave symposia.[2,3] The technological approach as well as the test results were described in these papers.

The work carried out under this portion of the contract brings the thin-film, etched-circuit rectenna to a high level of maturity, available for aircraft applications and serious consideration for space use. There are problems remaining, however. One is the radiation of harmonics. In addition the generation of spurious signals having to do with parametric oscillations in the rectenna have unexpectedly been found. This is the first known incidence of such parametric oscillations in rectennas after many years of development and application so that they are probably the function of the particular design of the rectenna. This is a phenomena that will need attention in future activity.

1.3 Evolution of Rectenna Technology that Provided the Foundation for the Thin-Film, Etched-Circuit Rectenna Concept

The thin-film, printed-circuit rectenna approach to be described in this report has evolved directly from the circuit format used in the conventional rectenna in response to a need for a lower cost, lighter weight, and more flexible rectenna that will operate efficiently at relatively low power levels. There have been several distinct steps in the technological evolution.[4,5]

The first step involved a transition from a three plane rectenna construction format as shown in Figure 1-1 to a two plane format shown in Figure 1-2 and 1-3 in which nearly all of the rectenna functions are carried out on the foreplane.[6] A physical realization of this is shown in Figure 1-4. The second step in the evolution involved redesigning the rectenna element to operate at a higher impedance level to retain good efficiency at relatively low incident microwave power densities that were felt to be desirable for rectennas in the upper atmosphere, in space, or at the edges of the rectenna for the solar power satellite concept.[7]

1.4 Origin of the Thin-Film Etched Circuit Concept

The factor that led directly to active work on the thin-film printed-circuit concept was the need expressed by personnel at the Wallops Flight Facility for a flexible rectenna that could be used at an altitude of 70,000 feet and at reasonably low power density levels on a balloon. The latter requirement suggested the use of the electrical circuit previously developed in a bar type construction format as the electrical circuit prototype for the thin-film printed-circuit rectenna.[1,6]

The transformation of the mechanical design of the rectenna element from the bar-type format to the thin-film format represents a different and perhaps unique approach to printed circuit design. The foreplane is a balanced circuit that does not use a ground plane (the reflecting plane, located a quarter wavelength behind the foreplane, is not a ground plane in the "slot-line" sense). The mechanical design of the thin-film format seeks to simplify

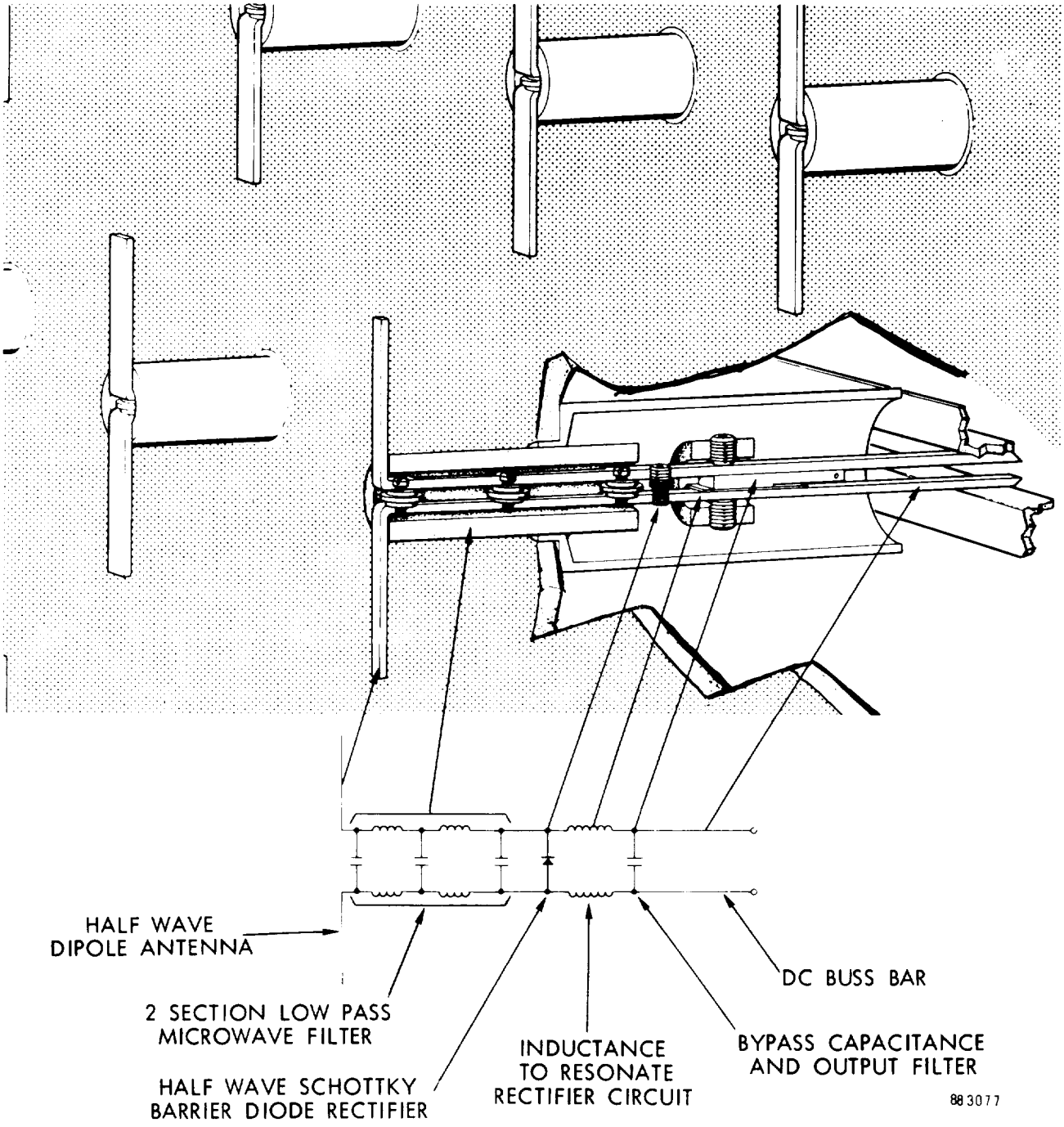


Figure 1-1. The Three Plane Rectenna Construction Consisted of (1) the Plane of the Half Wave Dipoles, (2) the Plane of the Reflecting Surface, and (3) the Plane of the DC Bussing Function. The Filtering and Rectification Functions of the Elements Ran Transverse to these Planes.

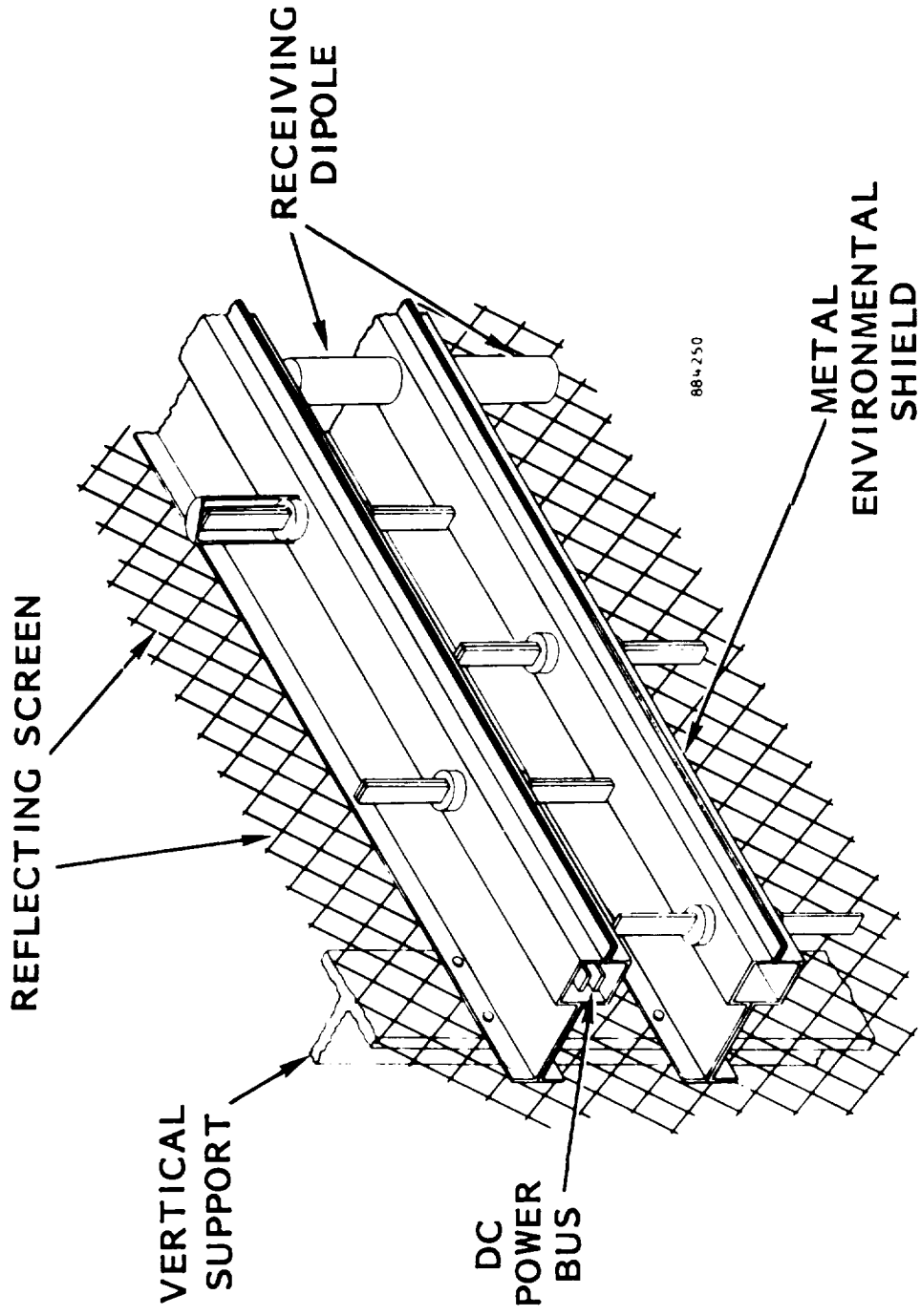
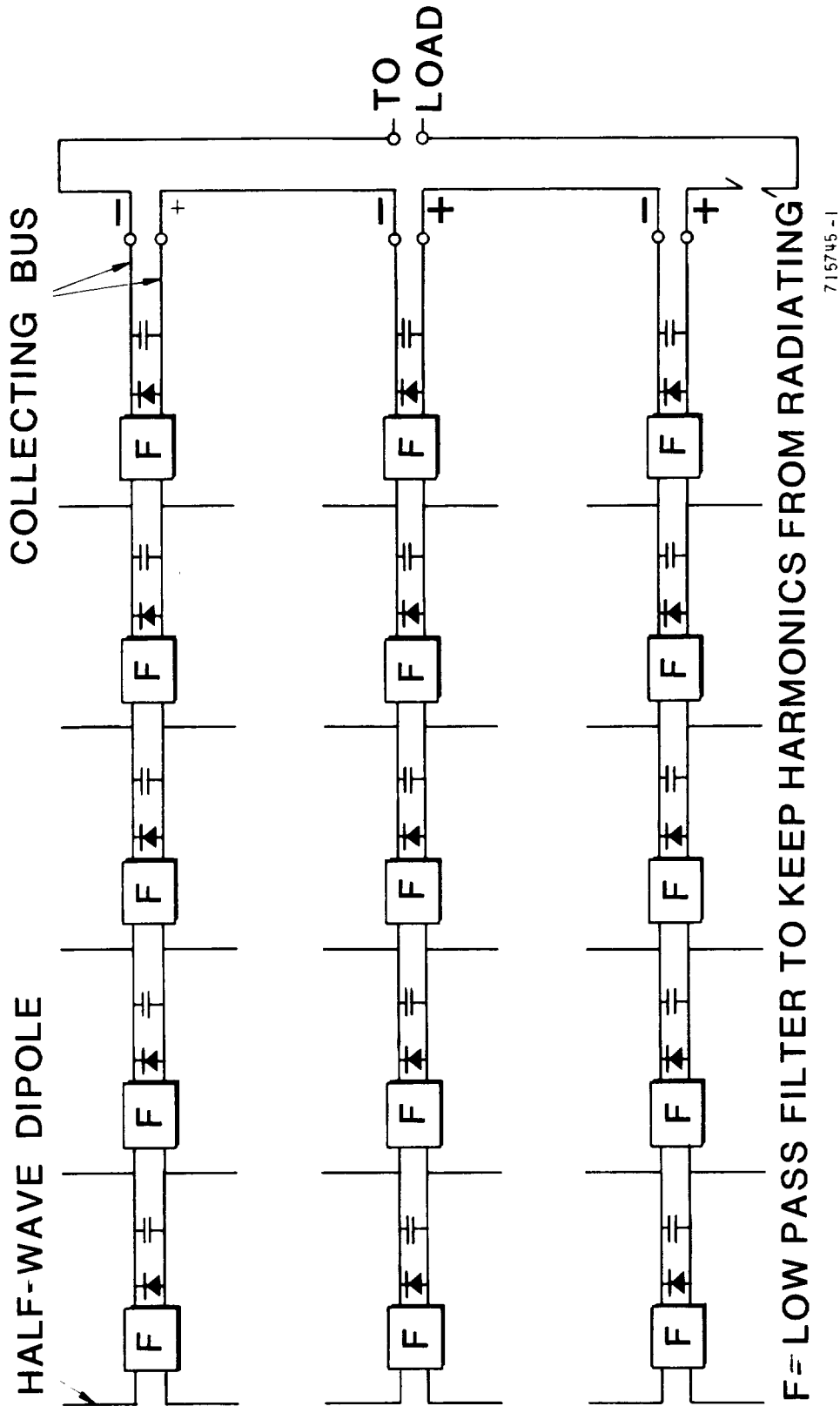


Figure 1-2. The Reorientation of the Rectenna Elements into a Two Plane Construction Format Grew out of the Necessity for a More Economical Construction Approach for the Solar Power Satellite Concept.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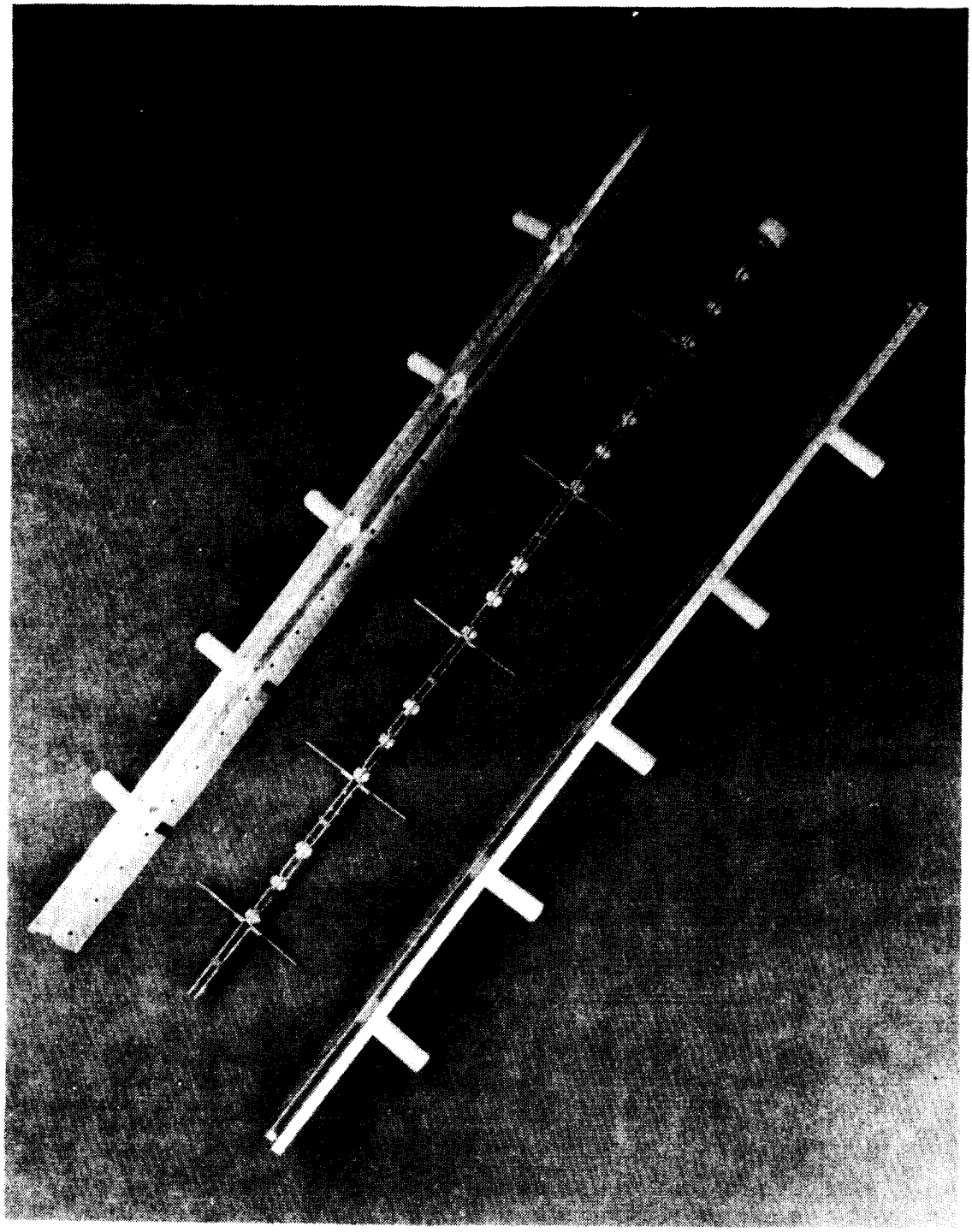

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Figure 1-4. The Item of Interest in this Photograph is the Connection of the Bar-Type Rectenna Elements into a Linear, Planar Structure, to Function as the Foreplane of the Two Plane Structure. This Format was Checked Out Electrically in Considerable Detail During a Previous Study. Shown also are the Environmental Shields, Not Needed for Space or High Altitude Atmospheric Platform Applications of the Rectenna.

fabrication by the elimination of any connections such as plated feed-throughs between the microwave circuits on the two sides of the film. It does this by etching the dipole antenna, the inductive sections of the low pass filters, and the DC bus bars on the top surface of the thin film. On the other surface the copper is etched to leave just enough material to form the capacitors associated with the low pass filters and the DC blocking capacitor. Figure 1-5 illustrates this arrangement.

Using this approach, a thin-film, etched-circuit rectenna was fabricated from a sandwich material consisting of one mil mylar bonded with an adhesive to one ounce (1.5 mil) copper on both faces. The resulting product is shown in Figure 1-6.

The electrical tests made upon a rectenna element cut from the larger sheet and tested with the special fixture to be described later indicated a moderate level of success but its efficiency was considerably lower than expected. Mylar as a base material also had the deficiencies of a relatively low softening temperature and of being sensitive to deterioration from ultra violet light.

The work performed under this LeRC contract may be considered as a major step in the evolution of the thin-film, printed-circuit rectenna. It has concentrated upon making the rectenna element (and therefore the rectenna) more efficient, and upon the use of plastic materials better suited for space application than mylar. The effort has also involved additional diode development which will favorably impact the performance of the rectenna.

1.5 Investigation of Rectenna Design for Frequencies of 20 GHz and Greater

Tasks six and seven of the work statement deal with investigating the rectenna principle at much higher frequencies and examining in some detail the design of a rectenna at 20 GHz.

The pursuit of these tasks was necessarily based upon the rectenna technology that had been developed at 2.45 GHz. However, it was found that a rectenna designed by simply scaling the rectenna from its 2.45 GHz thin-film, etched-circuit format did not appear attractive at such high frequencies because of a host of problems. Perhaps the most serious one was the large number of rectenna elements per unit area, because their density scales as the square of the frequency. Even if small diodes could be constructed in large numbers economically, the problem of assembly and carrying heat away from the diodes remained.

Because of scaling problems having to do strictly with the scaling of the thin-film circuit the use of a substrate on which the circuits could be silk screened became attractive. The standard substrate for such silk screened films is alumina ceramic. But alumina is a fair, if not good, conductor of heat so that filling the space between the foreplane and reflecting planes with

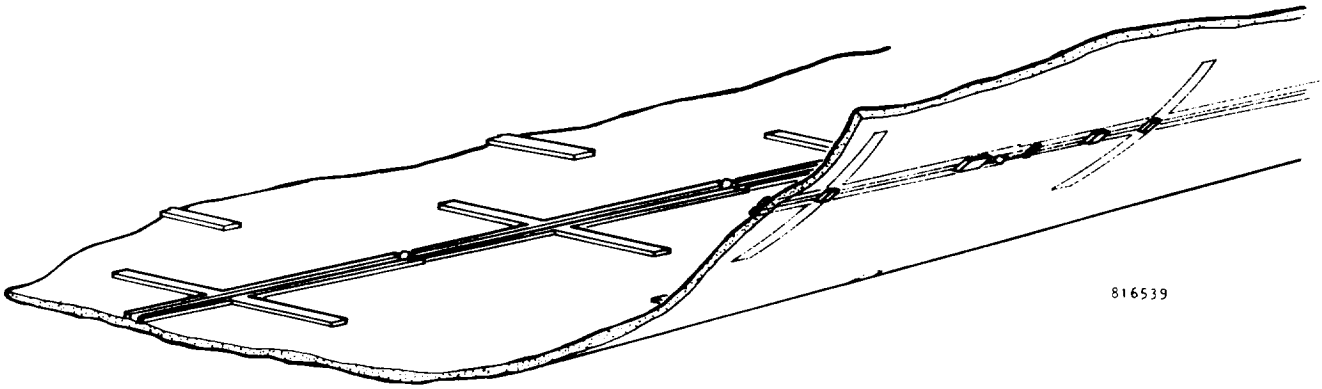

Figure 1-5a. Principle of the Thin-Film, Etched-Circuit Rectenna. Circuit Elements are Etched on Both Sides of Dielectric Film. There are No Interconnects between Etched Elements.

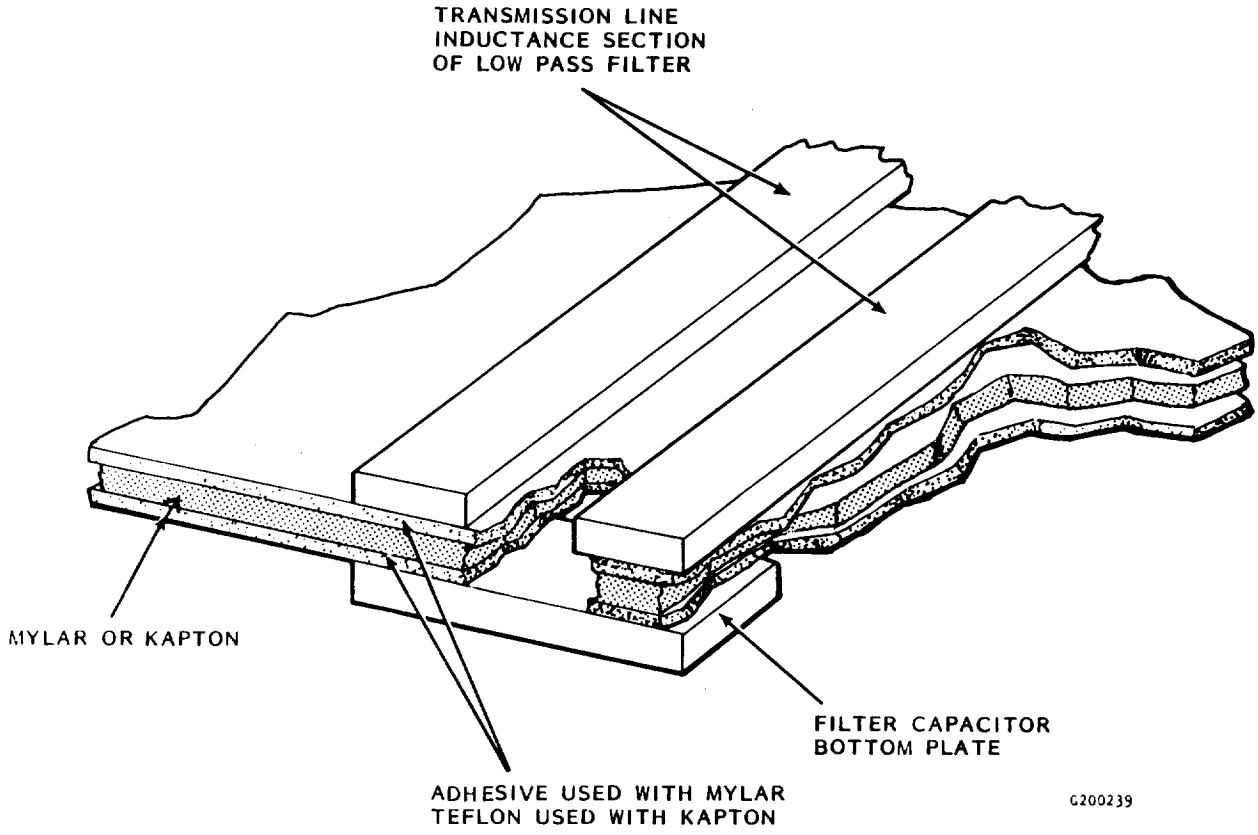

Figure 1-5b. Cutaway View of Rectenna Element Construction in Region of Capacitor for Low-Pass Filter Section. Low Dielectric Losses in the Film and Adhesive are Critical to High Efficiency. First Thin-Film Rectenna Used Mylar Dielectric and Adhesive Both of which have High Loss. Greatly Improved Rectenna Uses Kapton with Teflon as Adhesive.

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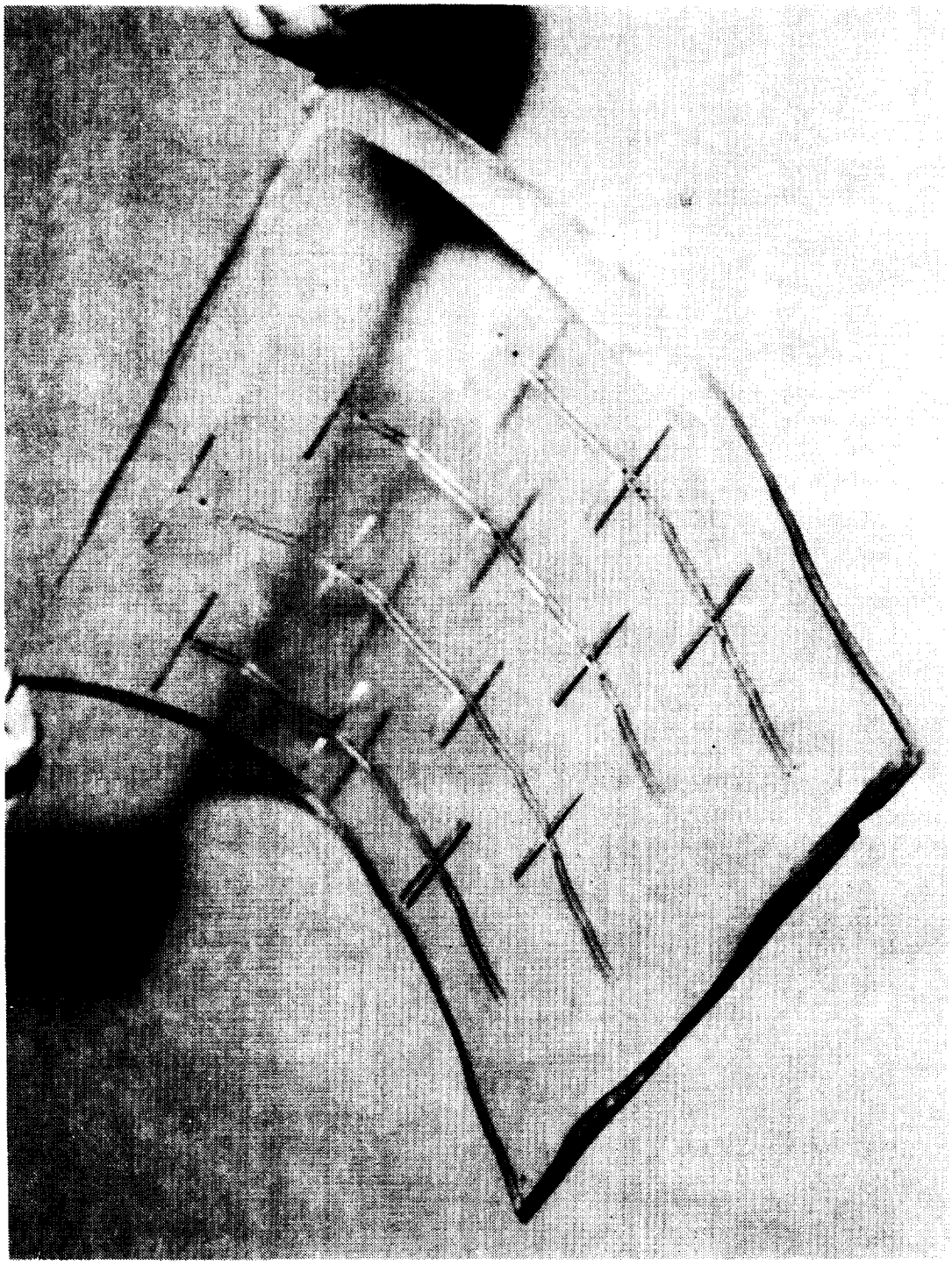

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Figure 1-6. Photographs of the First Thin-Film Printed-Circuit which Forms the Foreplane of the Rectenna. The Dielectric Substrate Used was Mylar which was Found to have Several Shortcomings as a Substrate for a Thin-Film Rectenna.

alumina ceramic would be an attractive solution to the problem of cooling an array. Cooling surfaced early as a problem because a high frequency rectenna must operate at a high power density to be efficient, and a combination of the higher density and the lower efficiency of a high frequency rectenna necessarily generates a large amount of waste heat per unit area.

Because alumina has a high dielectric constant of about nine, it also became evident that the half wave dipoles of the rectenna would become considerably shorter with the result that there would be more elements per unit area. But it was noted that any development involving a silk screened circuit on an alumina ceramic would be also applicable to a semiconductor substrate of silicon or GaAs which have about the same dielectric constant as alumina.

Over the time period that the contract was active, the technology of monolithic GaAs circuits had advanced so rapidly that at the time this report is being written the prospects of being able to build a monolithic rectenna on a wafer of GaAs became a reasonable possibility. For the future, a completely monolithic circuit is very attractive.

If the development of such a rectenna were begun now there would probably be a significant percentage of the diodes that would be inoperative, perhaps as high as ten percent, because of the imperfect nature of the GaAs wafer technology. However, the monolithic rectenna circuit could be so designed to tolerate the failure of such a percentage of the diodes. And over the near future it is expected that the quality of the GaAs substrate, which is a basic problem in much more complicated and currently much more important monolithic circuits, will be rapidly improved.

Ten years ago, there appeared to be no need for a high frequency rectenna because no high power cw transmitter technology existed at these frequencies. That situation has changed dramatically. The introduction of the gyrotron electron tube assures the availability of several hundreds of kilowatts of cw power at 20 to 35 GHz. And large mechanically-steerable 70 meter parabolic reflectors are being readied to be used with such tubes for deep space radar purposes. It is logical to expect that these breakthroughs in technology will be examined for their application to power transmission. The interest in rectennas for this application may be expected. It would be timely to consider undertaking monolithic rectenna developments at this time, even though there are still imperfections in GaAs technology.

2.0 MAIN TEXT: REPORT ON TECHNOLOGY PROGRESS BY TASKS

There were seven tasks to be performed under this contract. The report on technology progress will by task as outlined in the work statement of the contract with the exception of Task 5 which is the Review and Reporting Requirement. As indicated in the Introduction there were two main objectives, one associated with further development of the thin-film, etched-circuit rectenna at 2.45 GHz and the other associated with an extension of rectenna technology to much higher frequencies at 20 GHz and beyond.

The first four tasks are associated with the first general objective. Tasks 6 and 7 are associated with the second objective. The work statements associated with all of these tasks are presented below.

Task 1 - Single Element Printed Circuit Rectenna

The Contractor shall investigate and define the microwave and printed circuit design techniques that will be used to fabricate a single element printed circuit foreplane rectenna for evaluation. Using existing rectenna models, the Contractor shall establish electrical and mechanical requirements for the solid state component, and address thermal requirements for operation of the element in a vacuum environment.

The fabrication of the rectenna element shall include three recyclings of the artwork: the first to include the output bypass capacitor, and the second to fine tune the design, and the third to optimize the efficiency. Performance testing to determine such factors as efficiency, line and load characteristics, transient response, and standing wave ratios would be confined to a single element.

Task 2 - Design of a Complete Rectenna Array

The Contractor shall investigate the design of a combined foreplane, reflecting plane, and a separator. As part of the effort, methods of deploying a printed circuit rectenna array shall be considered. Mechanical, electrical, and thermal characteristics of various separator configurations shall be addressed.

After examining the various approaches, the Contractor shall select the "best" design and fabricate a small section of the combined array. This combination array shall consist of five foreplane structures without microwave diodes mounted on the substrates. No electrical testing will be performed on this combined array section.

Task 3 - Fabrication of a Large Area Sample Rectenna

Using the results of Tasks 1 and 2, the Contractor shall fabricate two sample rectenna structures less than 0.2 M^2 suitable for operation in vacuum.

This array shall be tested electrically in an ambient environment. After passing the electrical performance check, the sample rectenna array shall be sent to Lewis Research Center for operational tests in both ambient and vacuum environments.

Task 4 - Design of an RF Test Facility

The Contractor shall provide a suitable design for an rf test facility to be built at Lewis Research Center. The critical component requirements and performance specifications of the major rf excitation equipment and microwave measurement equipment shall be specified. As part of this task, the Contractor shall recommend rf components and advise the NASA Project Manager during the fabrication of the Lewis microwave test facility.

Task 6 - 20 GHz Printed Circuit Rectenna Study

The Contractor shall investigate the feasibility of developing a rectenna element capable of performing in the K-band region. This order of magnitude increase in operating frequency will allow the physical size of the rectenna to be reduced accordingly.

The study shall include, but not be restricted to, the effects of dielectric loss, semiconductor performance, diode technology, critical design parameters, and photoetching/layout techniques.

Task 7 - Preliminary K-Band Rectenna Design

The Contractor shall investigate and define the microwave and printed circuit design techniques that would be necessary to realize a K-band printed circuit foreplane rectenna.

This design will include the selection of an operating frequency/ substrate combination as to optimize the element's performance, specification of the proper rectifying diode, and preliminary artwork of the complete rectenna.

2.1 Single Element Printed Circuit Rectenna (Task 1)

This section will be organized by first establishing what the single rectenna element is and how it is tested. This will be followed by an account of finding the causes for the unexpected inefficiencies in the first rectenna elements made under NASA contract NAS 6-3006. This, in turn, will be followed by an account of developing a highly efficient rectenna element that is also greatly improved from the point of view of power handling capability and durability by a change in the film material.

2.1.1 The Single Rectenna Element and Test Procedures for It

Historically, rectenna development has proceeded by a test procedure that allows a detailed evaluation of a single rectenna element in a closed system that simulates the cell area that the element occupies in the rectenna. By this procedure, an accurate measurement of its efficiency can be made, and its impedance as seen by the incoming microwave beam closely approximated.

In the two plane format the rectenna element consists of a repetitive unit of the rectenna foreplane as shown, for example, in Figure 1-3 together with the metallic reflecting plane which is positioned about one quarter wavelength behind the foreplane section. The repetitive unit of the rectenna foreplane (Figure 1-3) consists of halfwave dipole antenna that couples to the incoming microwave beam or to incident microwave power in the special test fixture to be described later. The power from the antenna flows into a two section low pass filter which serves the dual function of energy storage for the half wave rectifier which follows it, and to attenuate the flow of harmonic power from the rectifier to the antenna. The rectifier is shunted across the output terminals of the low pass filter and its capacitance resonated out by a short section of transmission line which is terminated by a large bypass capacitor. The capacitor serves both as an effective short circuit termination of the transmission line and as a capacitor filter to minimize any microwave ripple on the DC power output of the element. The DC power output is collected on the two conductor strips that connect one element with another. The conductor strips serve a dual function of collecting the DC power as well as functioning as inductive sections of microwave transmission lines within the rectenna element itself.

The properties and performance of the rectenna element as just described is presented in great detail in References 1 and 6, including detailed mathematical modeling and computer simulation of performance that provides information on current and voltage waveforms, efficiency, harmonic content, etc. Reference 1 provides design procedures for the filter sections. It will not be the purpose of this report to go into similar detail.

For testing purposes the rectenna is mounted on a hinged door as shown in Figure 2-1. The door becomes part of an expanded waveguide test fixture shown in Figure 2-2. In turn the fixture becomes part of a closed measurement

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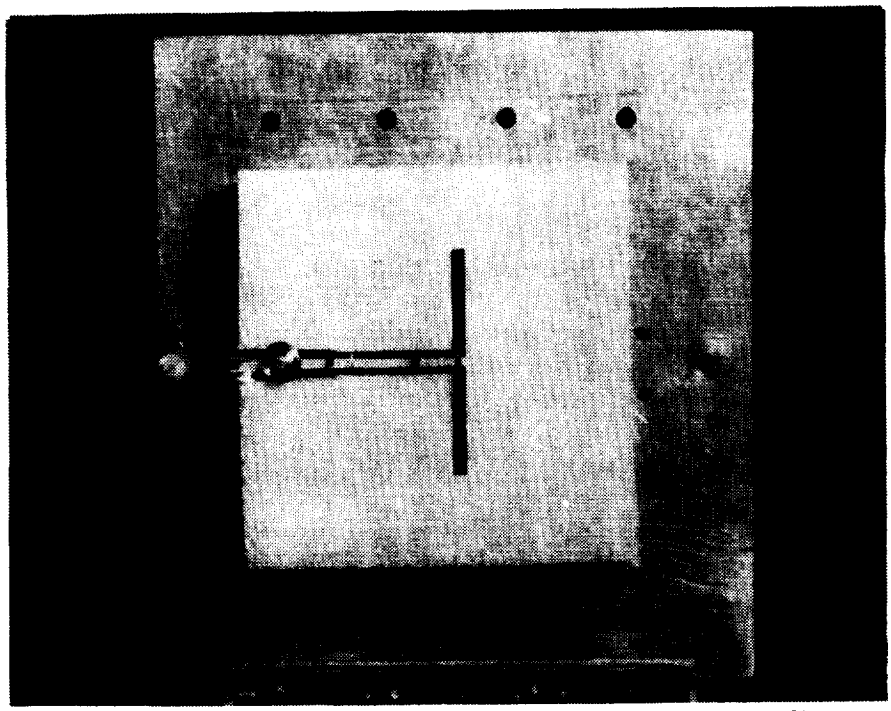

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Figure 2-1. Thin-Film Printed Circuit Element Shown in Test Position Mounted on a Hinged Door and Ready for Test.

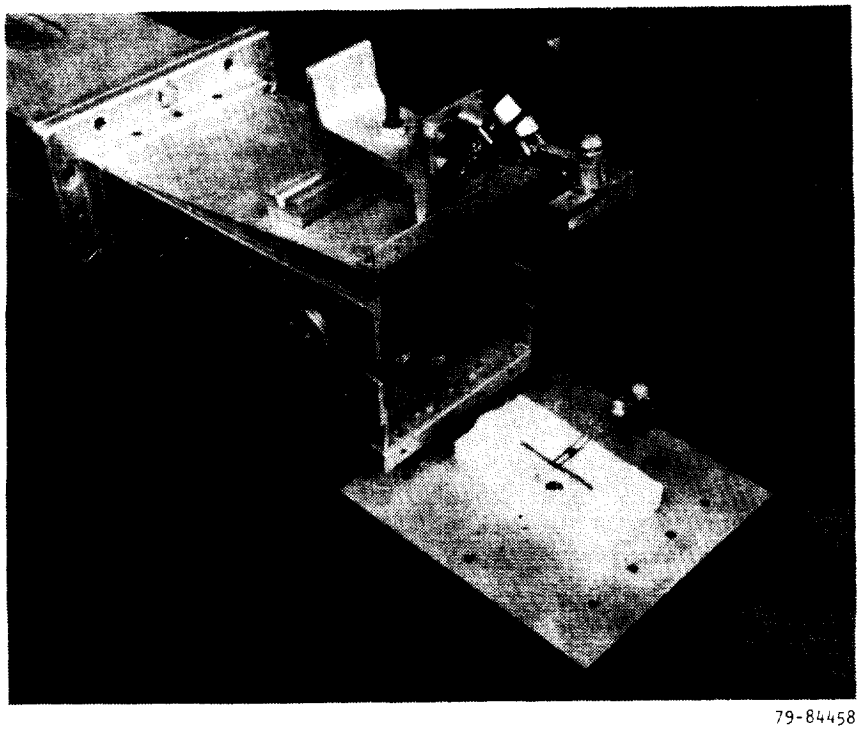

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Figure 2-2. Test Fixture for Testing an Individual Rectenna Element. During Test Door is Closed to Constitute a Closed System Check of the Rectenna Element as Shown in Figure 2-4.

system shown in Figure 2-3 in which accurate measurements of incident microwave power, reflected microwave power, and DC power output can be made.

Efficiencies can be computed from the measurements of the parameters shown in Figure 2-4. Efficiency may be stated in terms of overall efficiency defined as the ratio of DC power output to incident microwave power, or, electronic efficiency defined in terms of the ratio of DC power output to the microwave power absorbed in the rectenna element.

Measurements are also frequently made of the impact that DC load resistance and the microwave power input level have upon the input impedance level to the rectenna as measured at or near the plane of the foreplane. This information can be plotted on a Smith Chart.

Special attention is given to the validity and accuracy of the measurements. An effort is made to calibrate the microwave input accurately. This includes the elimination of the impact of harmonics in the calibration procedure by the use of low pass filters placed between the microwave generator and the system to be calibrated. Harmonic filters are also placed to eliminate the impact of harmonics generated by the rectenna element upon both the directional coupler with its power meter and the standing wave detector. Details of the system are given in Reference 6.

2.1.2 Determination of the Source of Inefficiency in the Original Thin-Film Etched-Circuit Rectenna Element

The first thin-film, etched-circuit rectenna element was modeled from a bar type rectenna element shown in Figure 2-5. Data on this element is given on pages 56 and 57 of Reference 6. The performance of this bar type rectenna with respect to efficiency as a function of microwave power absorbed is shown in Figure 2-6. Reflected power was so low that efficiency as function of incident power would have been nearly as high. The corresponding performance of the thin-film, etched-circuit rectenna element developed under the Wallops Flight Facility Support (Reference 1) is shown in Figure 2-7. There was a substantial difference in efficiency amounting to 13% between the anticipated efficiency and that which was experimentally measured.

Normally, a difference in efficiency of 13% could be tolerated. However, in the solid bar model case the inefficiency was 16% while in the first thin-film model it was 29% or almost twice as great. The power handling capability of the rectenna element will be determined by the power it must dissipate and dissipation must be by means of radiation alone if the element is in the vacuum of space. The application for the thin-film rectenna as stated in this study was to be in space. Moreover, the melting temperature of mylar from which the first element was made is relatively low, making dissipation an even more critical factor.

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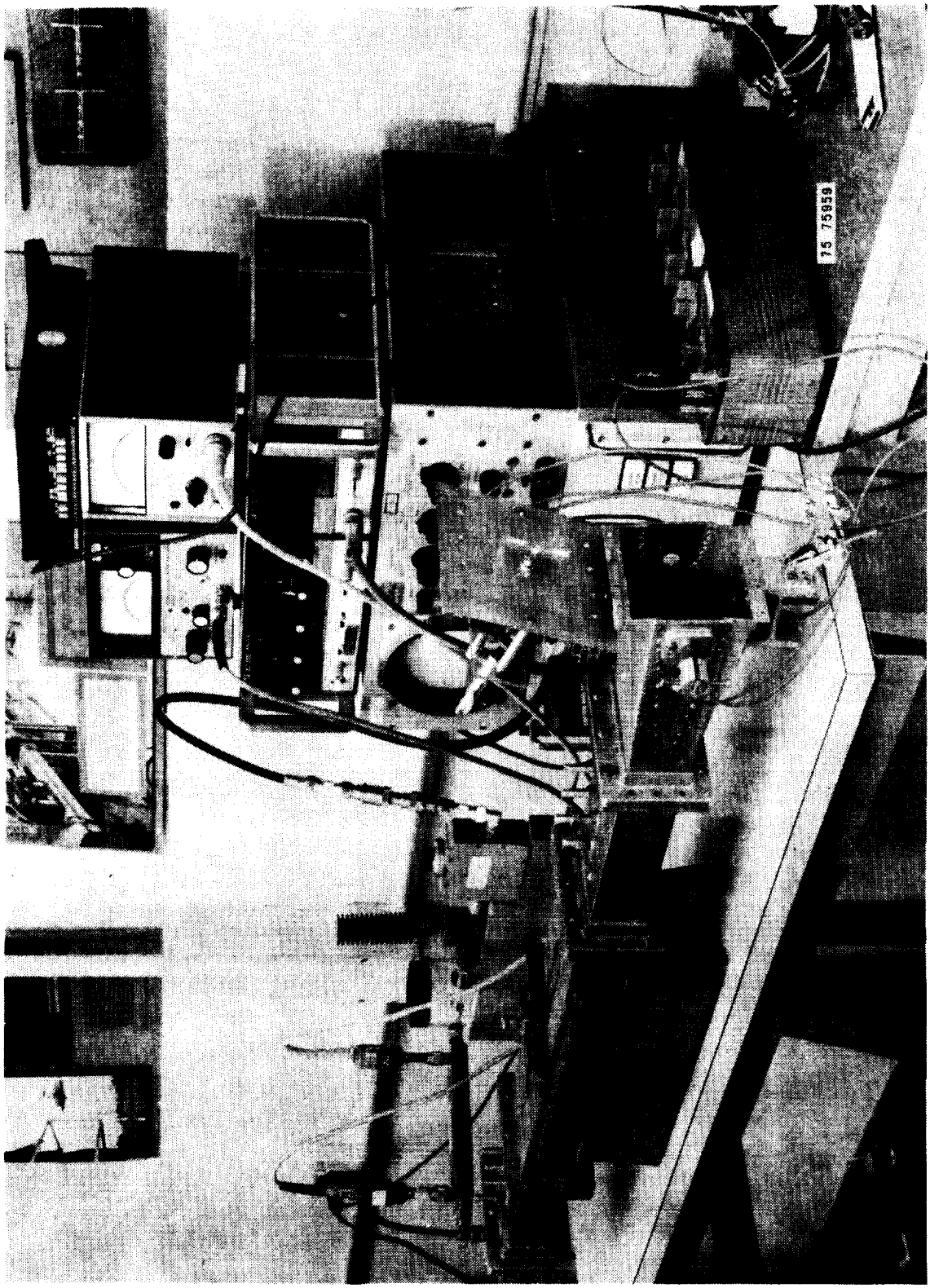

Figure 2-3. The Complete Rectenna Element Test Arrangement Utilizing the Expanded Waveguide Fixture.

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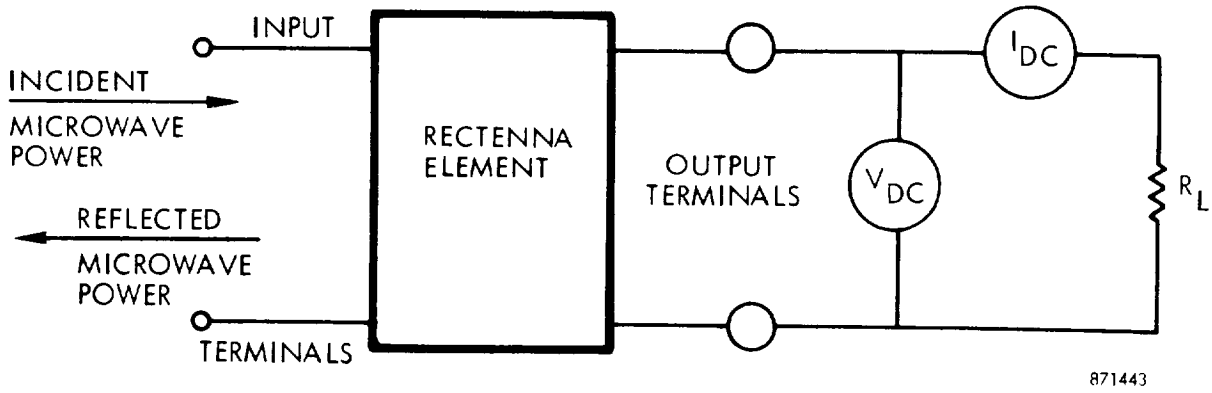

Figure 2-4. Input-Output Characterization of the Rectenna Element Showing the Measurements that can be Made to Compute the Efficiency.

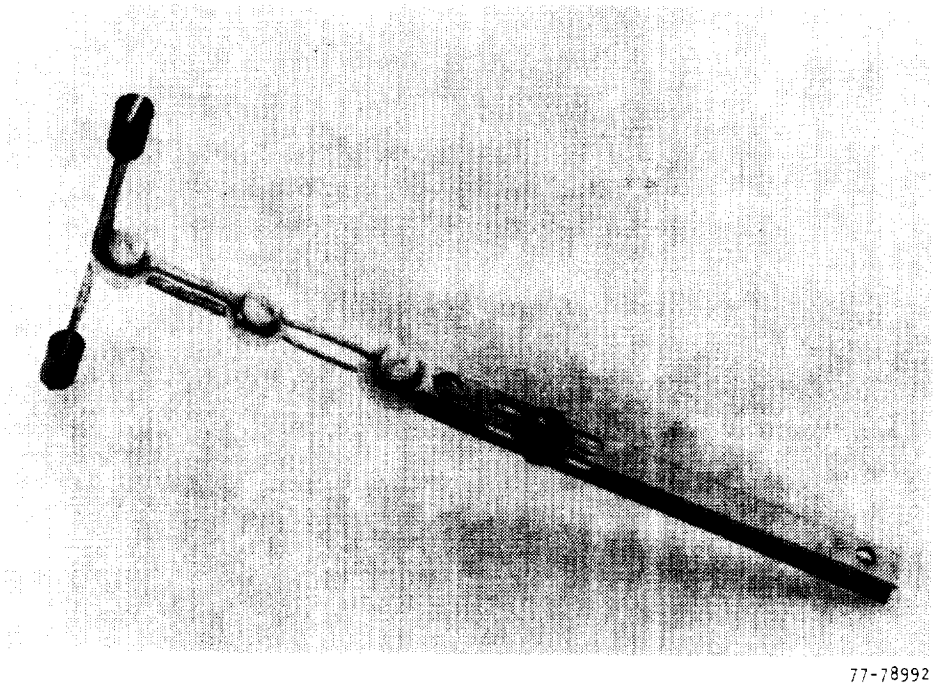

Figure 2-5. Special High Impedance Bar Type Rectenna Element that Served as a Model for the Circuit Design of the Thin-Film Printed Circuit Element.

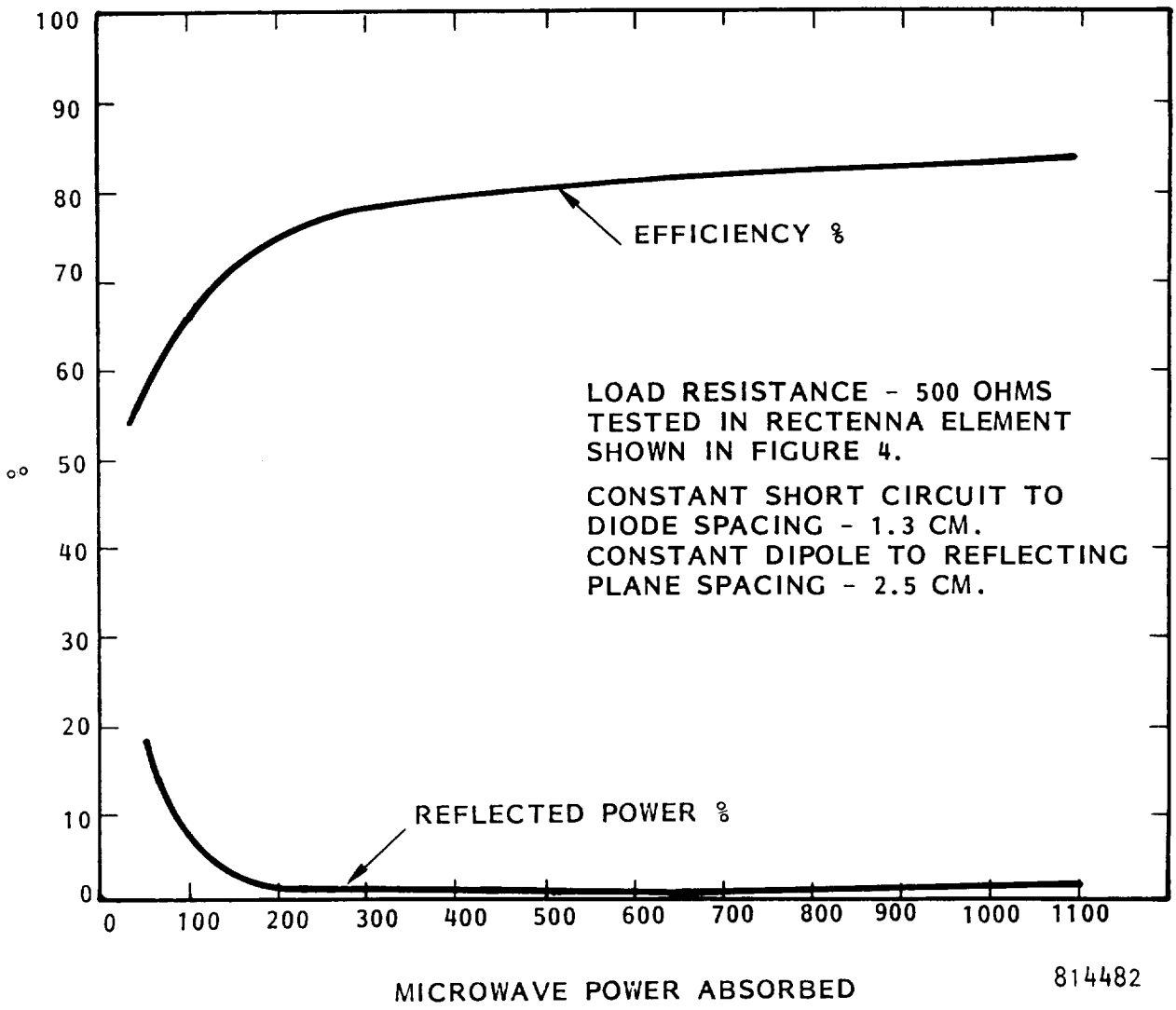

Figure 2-6. Efficiency and Reflected Power as % of Absorbed Microwave Power for the Combination of Specially Designed Microwave Diode and the Rectenna Circuit of Figure 2-5.

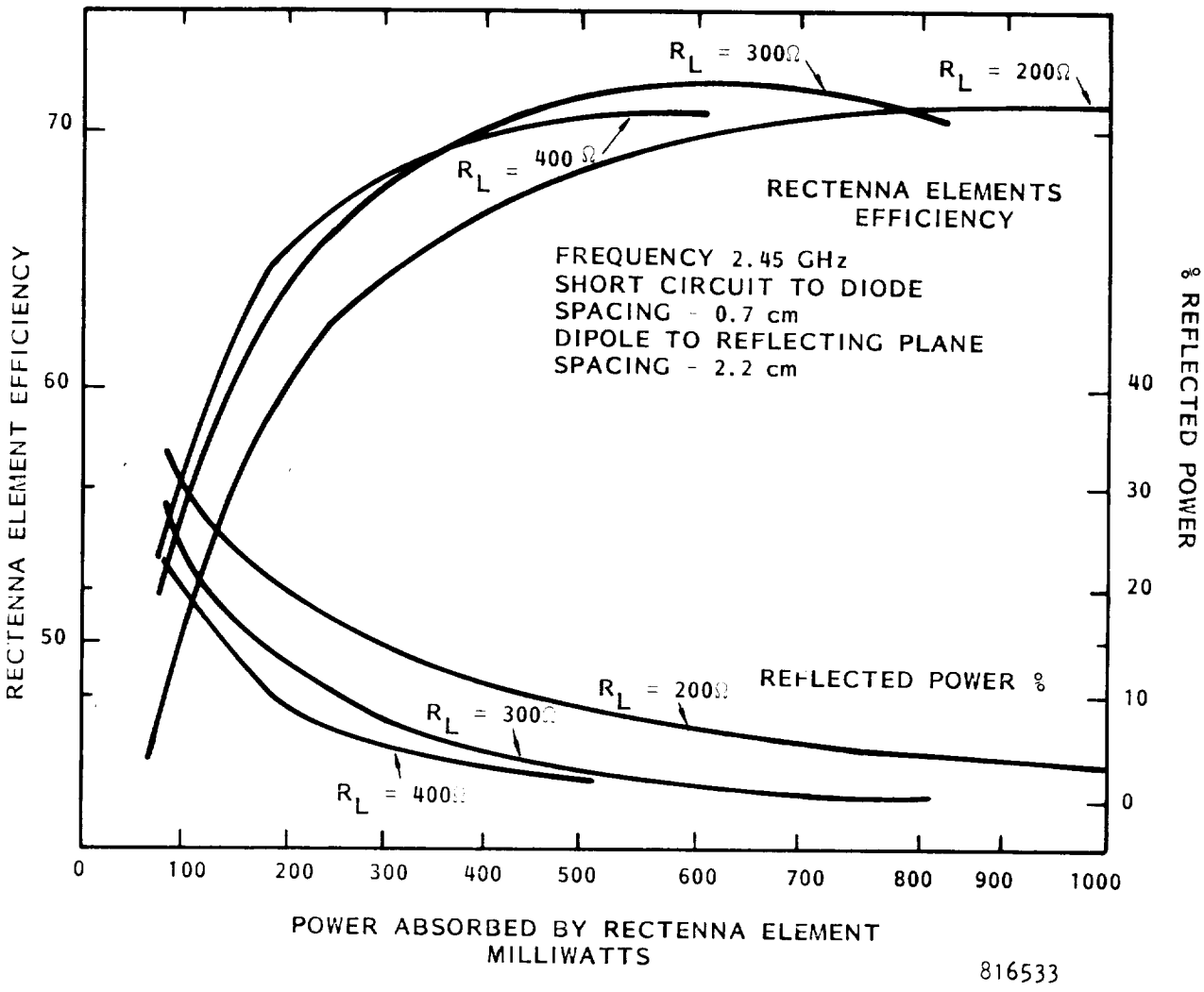

Figure 2-7. Test Results on Thin-Film Rectenna Element with Microwave Circuit of Bar-Type Rectenna Element Shown in Figure 2-5. Efficiency is Considerably Lower than that of Bar-Type Configuration.

Therefore, the first concern was with determining the source of the inefficiency, and then with correcting and improving the design, if possible.

Because the data for the bar type element and film type element were obtained using different diodes from a production lot of diodes, the first step under the new contract was to eliminate the diode as a variable by using the same diode in both the bar and film type elements. The resulting data was nearly identical, so that the diode could be eliminated as the cause.

After eliminating the diode as a potential source of inefficiency, the early effort under this contract was devoted to making measurements on the rest of the rectenna element design to determine the source of the excessive losses. These measurements involve making special test fixtures.

The first special test fixture that was made is shown in Figure 2-8, and shown again as it was inserted into the Hewlett Packard network analyzer in Figure 2-9. Basically the test fixture adapted the balanced configuration of the rectenna element to the unbalanced configuration of the coaxial connection to the network analyzer by splitting the balanced structure in half and holding one half above the ground plane which was a continuation of the outer conductor of the coaxial connection. The capacitor patches on the bottom of the element (see Figure 1-5) are clamped to the ground plane, as they should be electrically, and so adequately support the rectenna element mechanically. The characteristic impedance of the rectenna element is cut in half by this arrangement and comes close to matching the 50 ohm impedance of the network analyzer. The phase shift versus frequency properties of the rectenna element are not impacted.

The measurements of reflected power and transmitted power as a function of frequency made with this arrangement indicated poor transmission both because of resistive losses and because the cutoff frequency of the rectenna network was lower than anticipated from the electrical design. The resistive attenuation losses were greater than were anticipated from assumed properties of the dielectric material in the capacitors which was the only suspected source of loss.

To directly investigate the dielectric loss a technique for making direct measurements of capacitance and loss tangent on the capacitor used in the microwave circuit was developed. This involved the use of a holder, shown in Figure 2-10, for a sample capacitor whose area and dielectric thickness were carefully measured. Dielectric constants and dielectric losses could then be computed from the standing wave measurements made in a coaxial standing wave detector with a moveable probe. Very useful information was obtained from these measurements. Although we had expected the possibility of larger losses in the capacitor because of the unknown losses in the adhesive that was used to join the copper to the mylar, we were surprised to find: (1) the values of capacitances to be 30% greater than the design value, and (2) that the loss tangent in the mylar itself without the adhesive was .0056 or about three times that of the value of 0.002 which had been used to calculate losses in the capacitors in the low pass filters.

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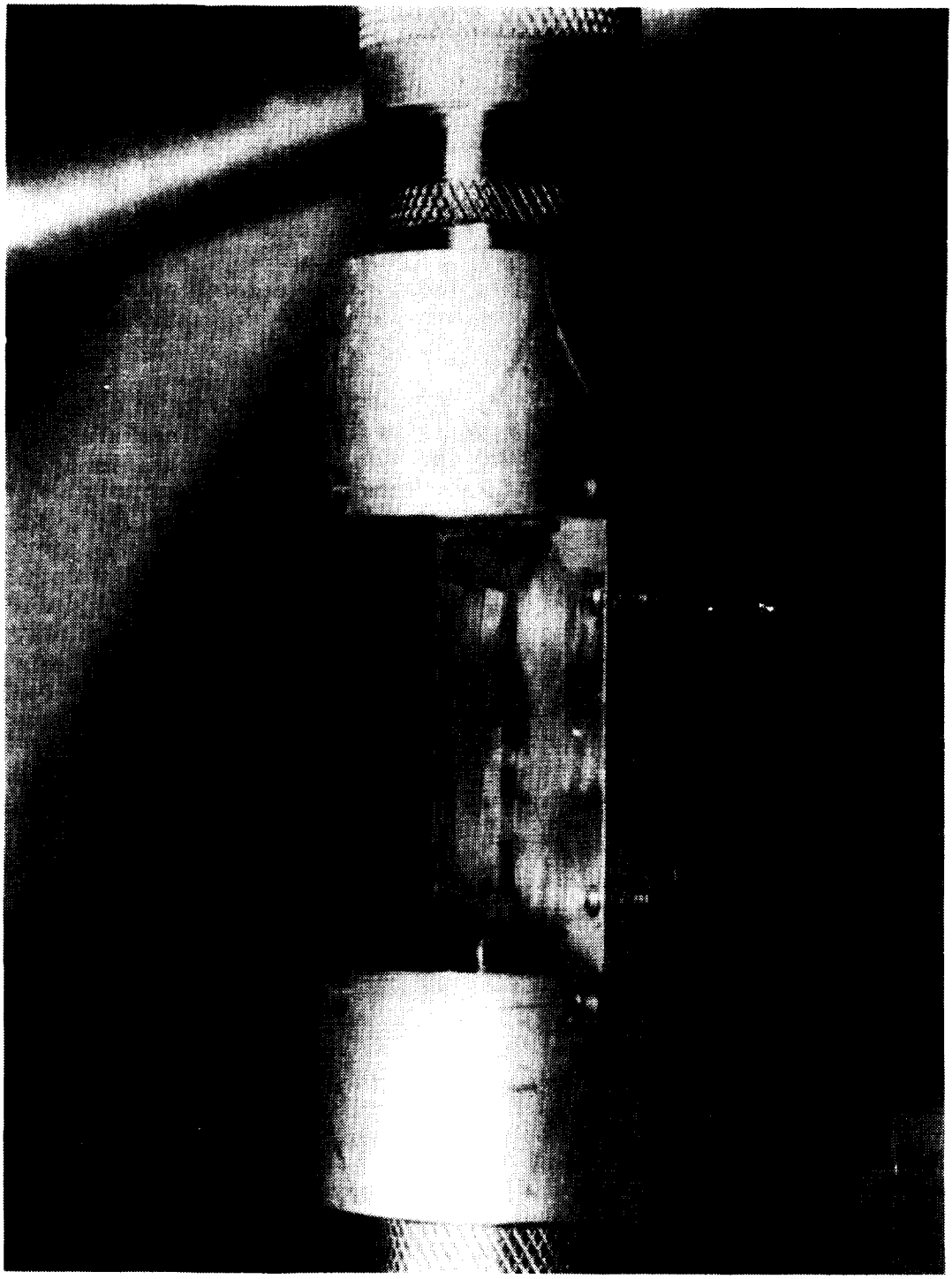

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Figure 2-8. Test Fixture for Determining Reflection and Transmission Properties of a Thin-Film Rectenna Element as a Function of Frequency.

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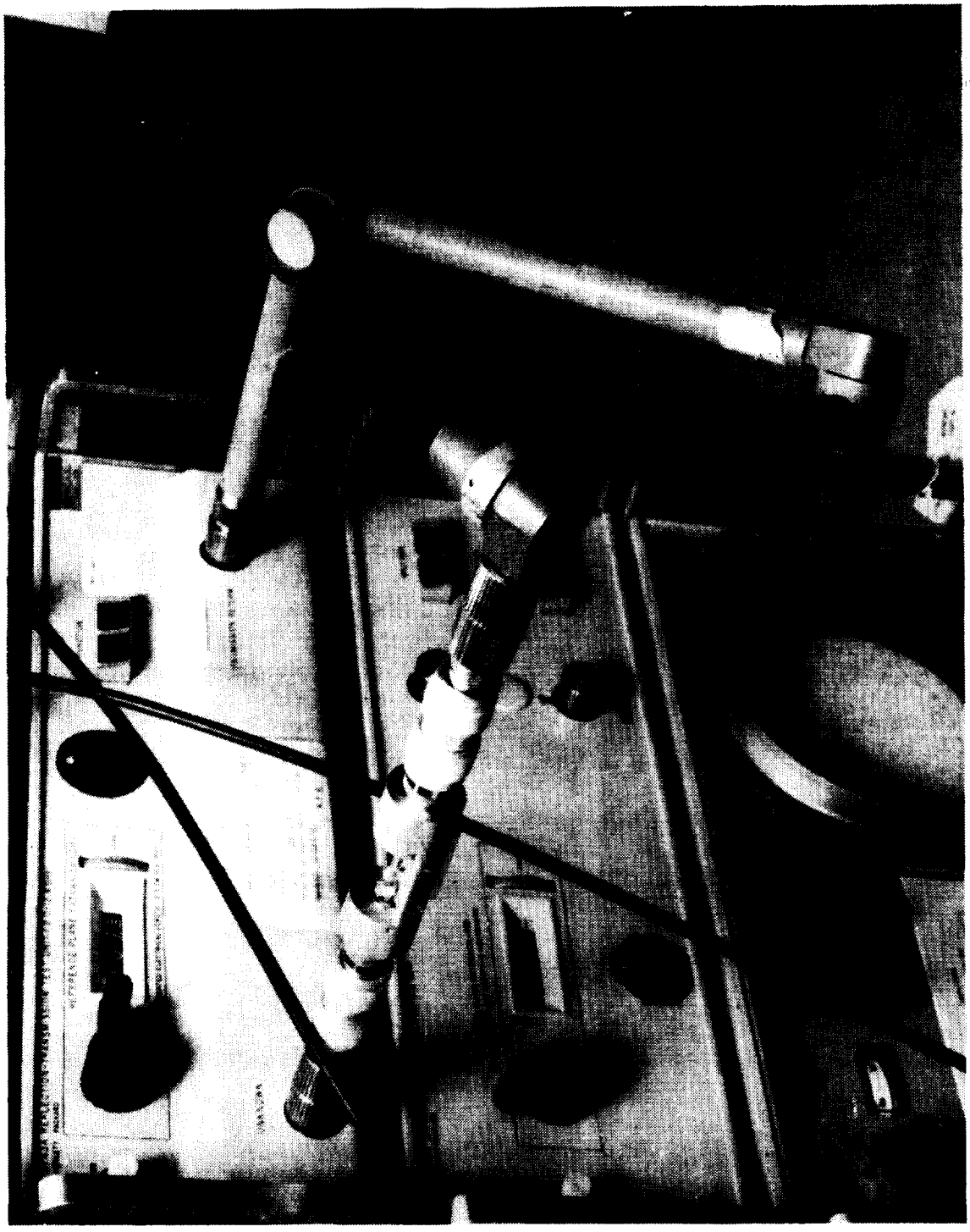

Figure 2-9. Test Fixture of Figure 2-8 Shown in Position in Hewlett Packard Network Analyzer.

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Figure 2-10. Special Capsule for Mounting Specimens of Capacitors Composed of Different Dielectric Materials. The Capsule Terminates a Sensitive Standing Wave Detector that could Resolve a Loss Tangent to Better than the Fourth Decimal Place, or the Difference Between 0.0026 and 0.0027, for Example.

The design of the rectenna element had been based upon data on mylar taken from Figure 3-15 on page 178 from the book "Dielectric Materials and Applications", authored by Von Hippel and published by the Technical Press of John Wiley. This is considered to be the authoritative text on the subject. From this graph the dielectric constant was read to be 2.2 and the loss tangent to be 0.002 at 2.45 GHz.

The matter of conflicting data was resolved by a conversation with William Westphal of the M.I.T. Insulation Laboratory who gave me values of loss tangent and dielectric constant from mylar at 2.45 GHz that were at substantial variance with those from Figure 3-15 in "Dielectric Materials and Applications". These values are listed on page 13 of Volume 5 of "Tables of Dielectric Material", Laboratory for Insulation Research, Technical Report 119. These tables could be obtained in Xerox form but are not as generally available as the book "Dielectric Materials and Applications". The dielectric constant and loss tangent are listed as a function of frequency in Table II of this report.

It was noted from Table II that the dielectric constant at 3 GHz (which is sufficiently close to 2.45 GHz) is 2.79 rather than 2.2. Hence the capacitors that were designed under the assumption of the incorrect value of 2.2 for the dielectric constant gave capacitances that were 2.79/2.2 or 1.268 times the design value. This mistake resulted in a lower cutoff frequency of the low pass filter than the design value, thereby introducing unwanted attenuation at 2.45 GHz.

The loss tangent for mylar at 3 GHz is given as 0.0061 in the tables. This compares with a value of approximately 0.0056 found from measurements on mylar itself.

The conclusion that can be made from this discussion is that there is good agreement between the experimentally measured values of loss tangent and dielectric constant in mylar at 2.45 GHz made at Raytheon and the data published in the tables for mylar. It follows that the data given in Figure 3-15 of "Dielectric Materials and Applications" is in error.

2.1.3 Establishment of Kapton F as a Suitable Dielectric Film Material

In the search for better film materials, Kapton coated with FEP Teflon emerged as a promising candidate. Here the real situation with respect to the published data and assumed loss in Kapton was reversed from that of Mylar. Kapton is particularly good in space with the low absorbed water content that it would have there. According to William Westphal of M.I.T., Kapton as received has a loss tangent of 0.008 at 3 GHz because of the absorbed water. However, after a bakeout at 100°C in air, the loss tangent improves to 0.0044, and after a higher temperature bakeout in vacuum the loss tangent improves to 0.0015, or four times better than mylar as we are now using it. Furthermore, Kapton will withstand a much higher operating temperature so that the operating temperature of the diode which can be as high as 200°C becomes the limiting factor in the

amount of the heat that can be radiated. The great improvement in the power handling capability of a rectenna in space made possible by changing to Kapton from Mylar is immediately evident.

Kapton, however, will not adhere tenaciously to copper cladding of the thickness that is needed and even when a thin copper film deposit is made in vacuum the subsequent addition of a thickness of electroplated copper results in an unsatisfactory product. The current technology, therefore, is to apply a thin adhesive to the Kapton before bonding copper to it, usually by compression at elevated temperatures. This adhesive is quite often Teflon. DuPont makes a commercially available product designated Type F Kapton with the Teflon already coated to the Kapton.

According to the DuPont Engineering Technical Service Type F Kapton bonds well to copper but we found there are no copper to Kapton F laminates commercially available because the bonding of the copper to the teflon requires high pressure at temperatures so high that an electrically heated press is necessary. Commercial laminates using copper bonded to Kapton use a bonding agent that will bond at the lower temperatures of steam heated presses. The low temperature bonding agents normally used are lossy at microwave frequencies.

Fortunately, we found suitable electrically heated presses within the Raytheon organization that could perform the bonding although the area of the press was limited which presented a limitation to the number of rectenna elements that could be incorporated into a single fabricated rectenna section.

The first laminates attempted with this equipment used only 0.1 mil of teflon film on the 1 mil Kapton core. The bonding was not satisfactory. The teflon thickness was increased to 0.5 mil on each face of the Kapton. These laminates were quite satisfactory and were used for subsequent work.

Measurements at 2.45 GHz were made of the loss tangent and the dielectric constant of the Kapton F material when bonded to copper. The loss tangent was established as 0.0033. This loss is considerably lower than the loss measurements of 0.0055 on Kapton alone and can be attributed to the presence of the FEP teflon which has a very low dielectric loss.

The measured value of the dielectric constant of the composite material was 3.09 or substantially higher than the computed dielectric constant of 2.63 for the composite material. This discrepancy may have been caused by improper compensation for the fringing fields surrounding the assumed one sixteenth inch diameter test capacitor or in a larger actual diameter of the capacitor than that of the mask.

2.1.4 Design of the Rectenna Element made from Kapton-F Copper Laminate

With a suitable laminate established, attention was given to the design of the rectenna element itself. There were three reiterations of the art work

that established the dimensions of the two masks that were necessary for etching the finished circuit elements on the two sides of the laminate. In addition to determining the details of the art work for the rectenna element it was necessary to make some assumptions about the spacing between rectenna elements. The assumption used was that the spacing would be the same as it was for the highly successful rectenna made from bar type rectenna elements.

The rectenna elements were spaced 7.772 cm (3.060 inches) apart in rows that were separated by 6.731 cm (2.650 inches). The 1:1 exact scale of the resulting art work for the rectenna elements is given in Figures 2-11 and 2-12. The dimensioning of the artwork may be obtained from Figure 2-13.

The electrical design of the thin-film rectenna has deviated some from that of the prototype bar-type element (Figure 2.5) in response to the difficulty of constructing mesa type diodes with zero bias capacitance, C_{to}, as low as one picofarad and in response to the experimental observation that a diode of 3.0 pfd can be positioned across the transmission line near the point of and in place of the inboard capacitor of the second section of the low-pass filter without substantially changing the performance. The physical capacitor is eliminated but is electrically replaced by a small portion of the effective capacitance of the diode when it is in operation. The larger area of the junction in the 3.0 pfd diode increases the power handling capability of the element.

The use of a higher capacitance diode implies a lower value of resistance of the dc load, which means that the voltage drop across the Schottky barrier diode becomes a more important contributor to inefficiency unless the DC power output of the element is increased to restore the same voltage across the DC load. However, it is important to note that as time has passed it has been recognized that the rectenna element, particularly in its Kapton F format will handle much more power in a vacuum environment than had been previously thought. At the same time it was found that the anticipated space and aircraft applications for the rectenna would need a fairly high power density level of from 200 to 1000 watts per square meter, or an equivalent output of from 1 to 5 watts from each of the rectenna elements. Therefore, the higher capacitance diode is desirable for space and aircraft application but not for application as an efficient element on the periphery of the ground based rectenna in a Solar Power Satellite transmission system for which the prototype element in Figure 2-5 was developed.

The electrical parameters of the final design are shown in Figure 2-13. The values of inductance for the π section equivalents of the two sections of transmission line are given at 2.45 GHz. The values of capacitance for the π section equivalent are merged with that of the patch capacitors to provide the total capacitance value shown. The procedures for deriving these values are given in Section 3.4 of reference (1).

The design calls for the second low pass filter to serve as an impedance transformer to raise the impedance level of the microwave rectification circuit. At 2.45 GHz it is designed for 90 degrees phase shift and serves as a quarter

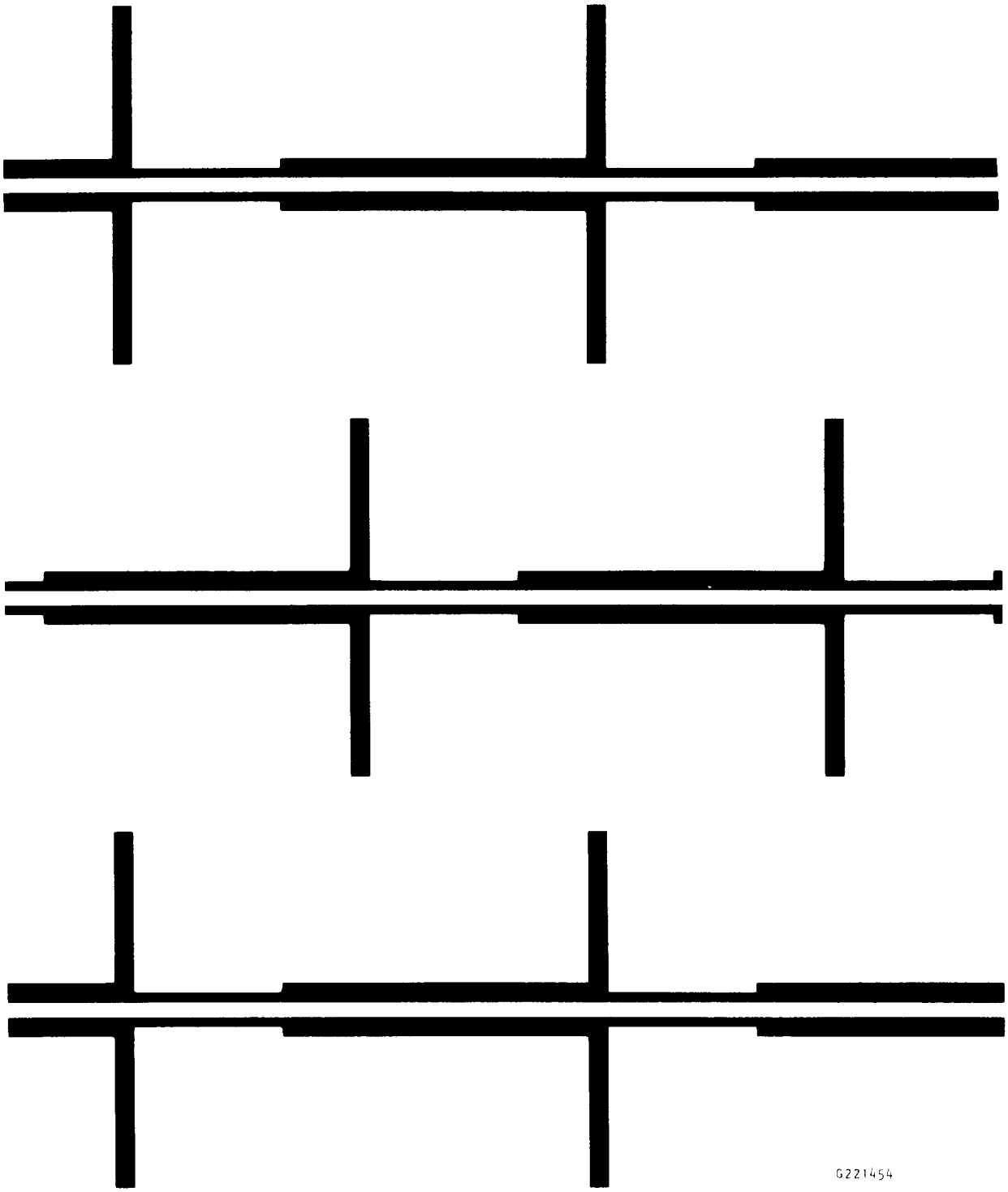

Figure 2-11. Exact Layout of Etched Circuits on Top Side of Thin-Film Rectenna. Refer to Figure 1-5.

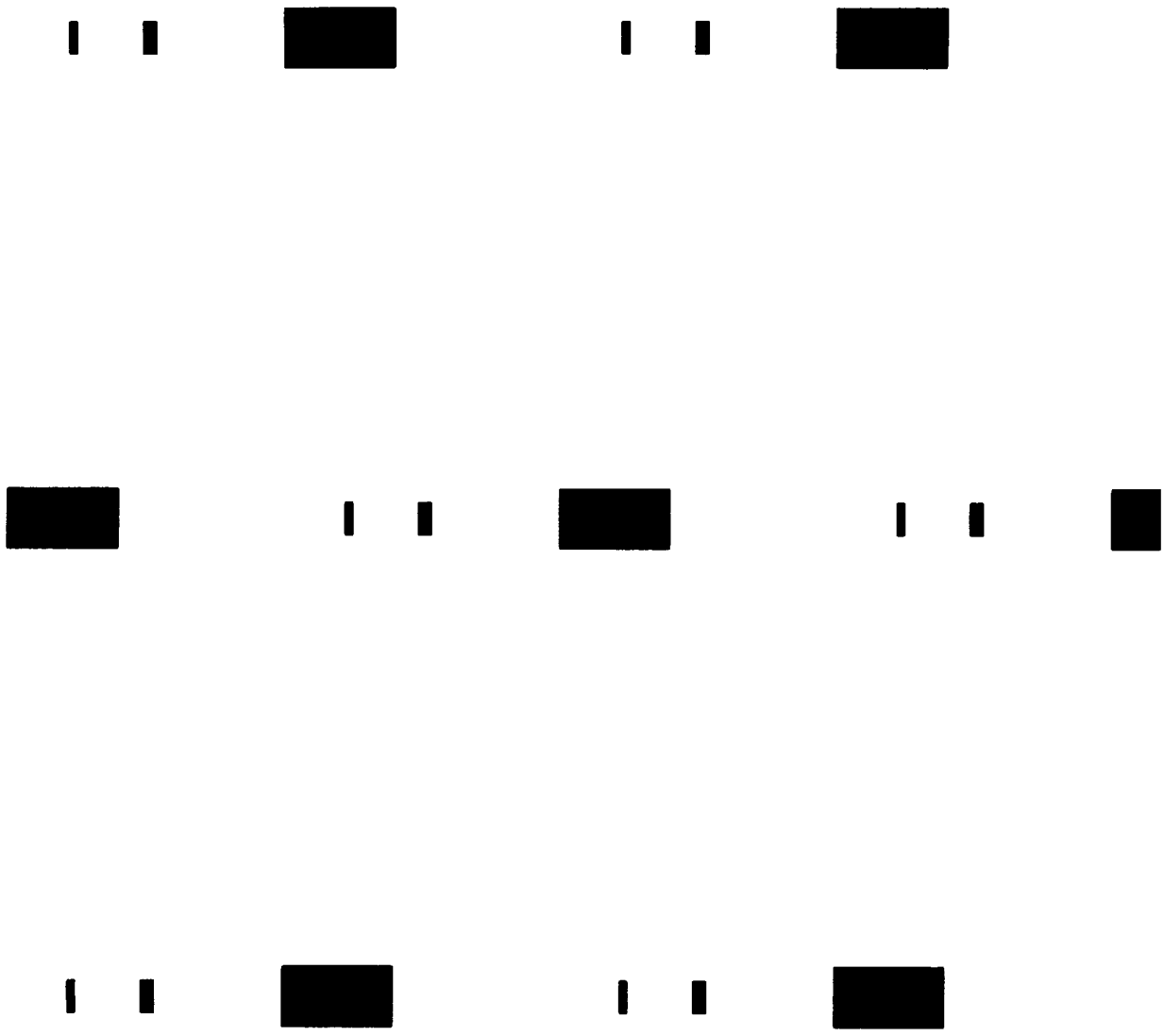

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Figure 2-12. Exact Layout of Etched Circuits on Bottom Side of Thin-Film Rectenna. Refer to Figure 1-5.

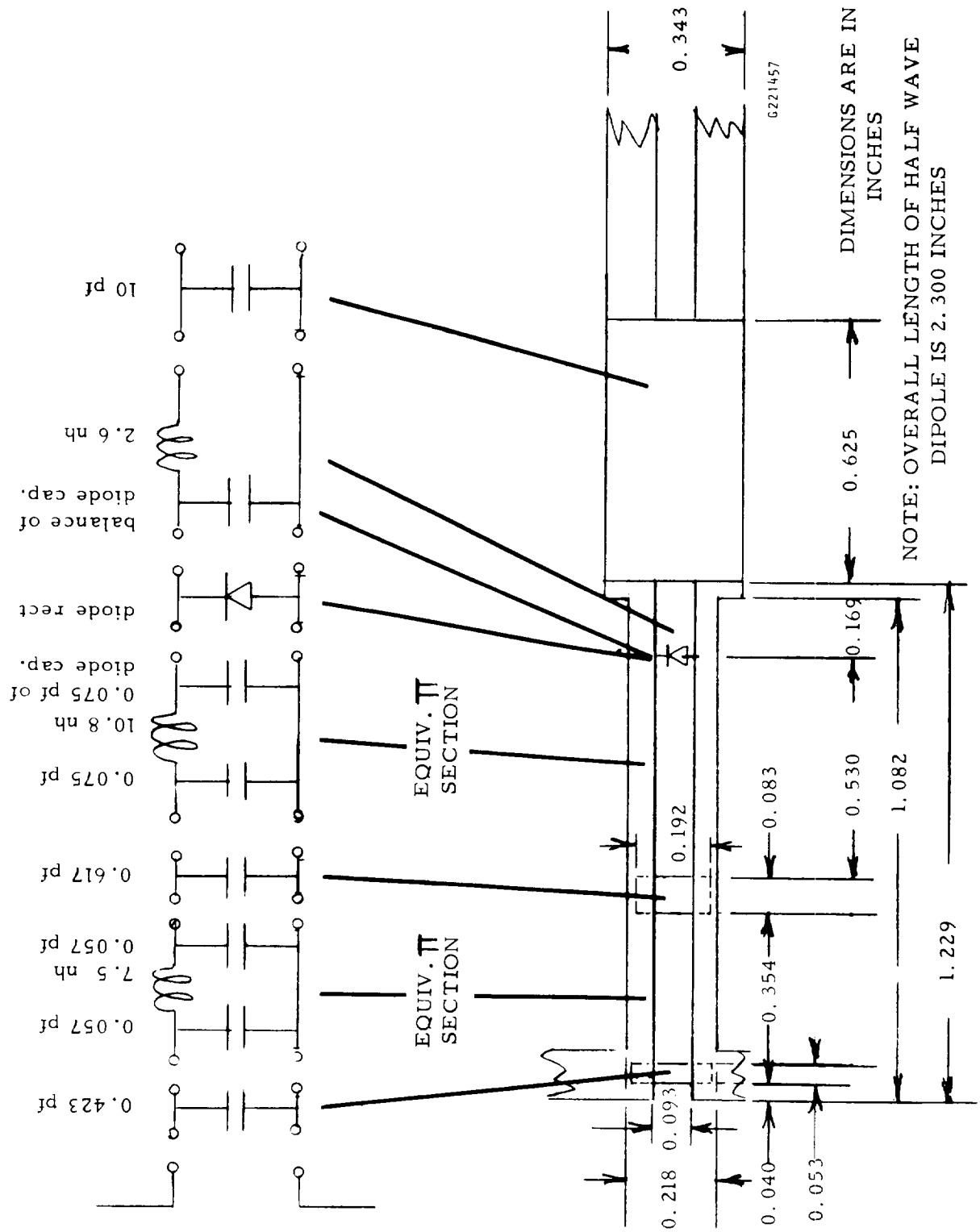

Figure 2-13. Physical and Electrical Design of Rectenna Element. At Bottom the Dimensions that Produce the Masks Shown in Figures 2-11 and 2-12 are Shown. At Top, the Electrical Schematic that Matches the Physical Design is Shown.

wavelength long line of 180 ohm characteristic impedance. The antenna terminal impedance is 120 ohms and the first low pass filter also has a characteristic impedance of 120 ohms. As seen by the rectifier circuit the impedance at 2.45 GHz looking toward the antenna is therefore 270 ohms. A reflectionless match therefore implies that the microwave impedance appearing at the rectifier circuit be 270 ohms. The DC load resistance that produces the match is typically 1.2 to 1.3 times the microwave impedance.

The data of Figure 2-14 indicates that the associated DC load resistor have a value in the range of 300 to 400 ohms to produce a good match to the rectenna. But such a value produces an excessively high peak inverse voltage for the current Schottky barrier diode when the DC power output exceeds 1.6 watts. Clearly, the rectenna element needs to be redesigned for lower impedance level when producing power levels of several watts.

2.1.5 Electrical Performance of the Rectenna Element in Its Final Configuration

Considerable data was taken on the rectenna element at each design reiteration. The data of greatest interest, however, is on the final design. The kind of data of greatest interest relates to efficiency and power handling capability. There are two different efficiencies of interest. One is the rectification efficiency of the rectenna element which is defined as the ratio of DC power output to the microwave power absorbed by the element. The power absorbed is equal to the incident microwave power minus the reflected power. The other efficiency is overall efficiency which is defined as the ratio of DC power output to the incident microwave power.

Figure 2-14 shows the overall efficiency as a function of the DC power output and the value of DC load resistance used. It also shows the reflected power as a function of DC power output and load resistance. The maximum DC power output is limited by the reverse voltage breakdown of the diode which is the sum of the DC output voltage plus the peak negative of the ac voltage waveform. This sum cannot be exceeded without a drop in efficiency. If the inefficiency is too great, the diode will be burned out. It is noted that a value of 200 ohms of load resistance has permitted a DC power output of 3.5 watts before the efficiency begins to droop. Later, in Section 2.3 when the power output of a rectenna array composed of these elements is discussed, DC power outputs in excess of 5 watts per rectenna element were obtained at even lower values of load resistance.

It will be noted from Figure 2-14 that there is considerable reflected power from the 200 ohm load. When the reflected power is subtracted from the incident power the rectification efficiency of the rectenna element is 84% at 3.5 watts and 85% at 2.5 watts of DC power output.

The rectification efficiency is probably the item of greatest interest when testing the individual rectenna element because it has been established that the collection efficiency in a properly designed rectenna can approach

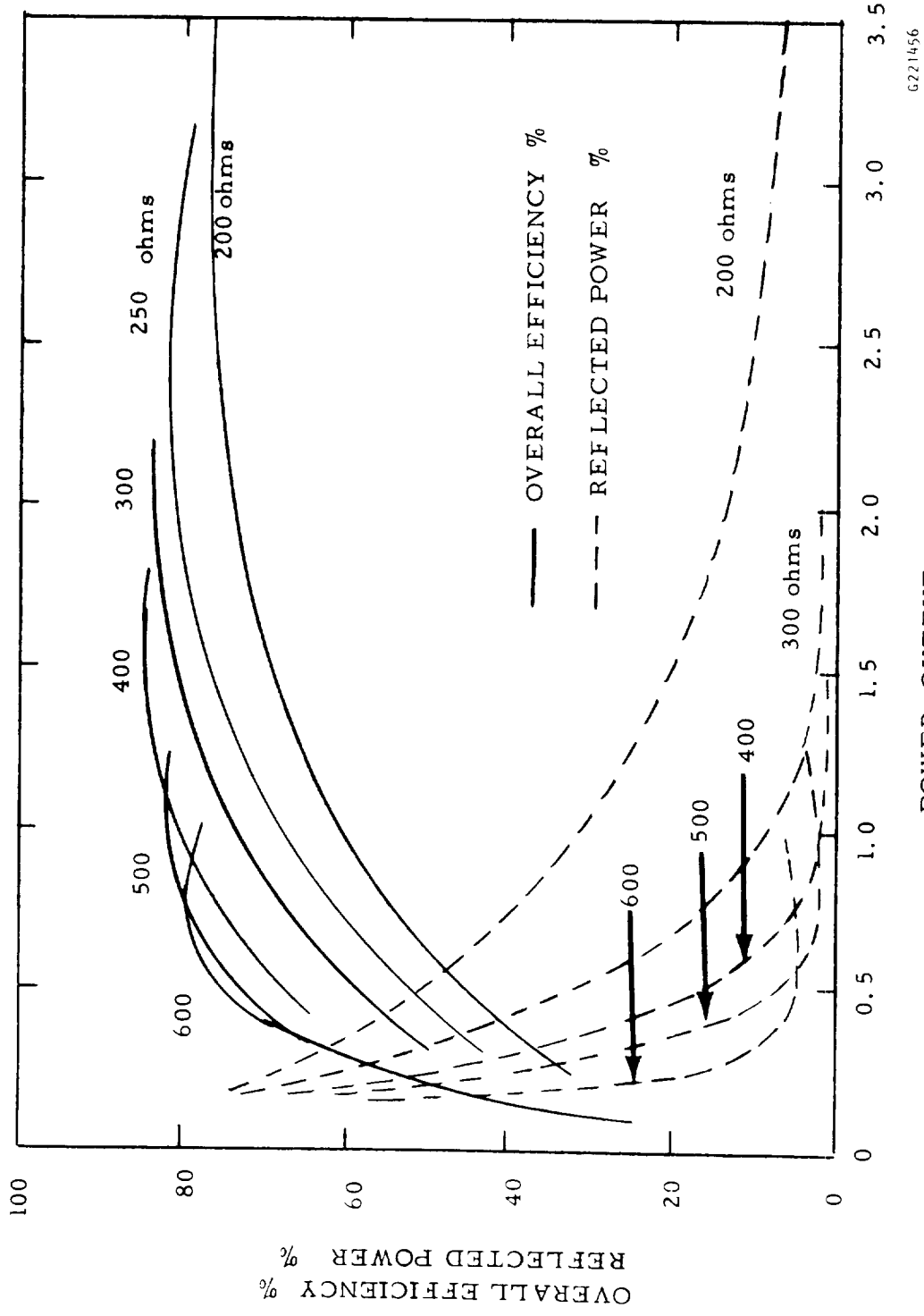

Figure 2-14. Overall Efficiency and Reflected Power Percentage are Shown as Functions of the DC Power Output and the DC Load Resistance. Power Output at Higher Values of DC Load Resistance is Limited by Start of Power Flow in Reverse Direction through the Diode. The Diode Absorbs this Power.

100%. One of the factors that must be taken into account in testing individual rectenna elements over a range of power input is that the effective capacitance of the diode is a function of operating power level so that for optimum performance the rectifier circuit should be retuned by repositioning the microwave shorting (bypass) capacitor. In a printed circuit format, however, this is not possible. The implication is that for optimum overall efficiency of the rectenna element, the operating power level must be specified.

There is considerably more test information on the power handling capability of the individual rectenna element along with that of arrays of the elements in Section 2.3.

2.1.6 Diode Development and Procurement

The Schottky barrier diode is an essential element in the rectenna. The use of GaAs as the base material is especially important in the rectenna application because of the high efficiency of diodes made from this material and because of its capability to withstand considerably higher operating temperatures than silicon. Because these diodes are not used for any other purpose, their development has always been part of the rectenna development effort.

The basic Schottky diode for the rectenna application was developed under contracts with MSFC and JPL in the 1972 and 1975 time period. However, the diode was packaged in a format not suitable for use in the thin-film printed-circuit rectenna. In the decision making process as to how the diode would be repackaged, the ultimate cost of the diode in large production volume was considered and it was decided to package it in a miniature glass package that is commonly used for mass production of all forms of diodes. This form shown in Figure 2-15 was used during the thin-film, printed-circuit, rectenna development for Wallops Flight Facility and during the early phases of the current LeRC activity. It worked out well and it was intended to use this for the entire LeRC activity. However, during a move of the Special Microwave Devices Operation from Waltham to Northborough the tools were lost to make these diodes. Because of the cost of replacing these tools, it was decided to repackage the diode in a ceramic pill package of the format shown in Figure 2-15. At the same time it was decided to increase to zero bias capacitance of the diodes to reduce the scrap involved in making the mesa type diodes.

The impact of this design change in packaging and the increase in capacitance upon the design of the rectenna element was considerable. For the rectenna element to perform satisfactorily it was found necessary to eliminate the inboard capacitor and reposition the diode as shown in Figure 2-13. In the final format shown in Figure 2-13 the performance actually appeared to be better than for the glass diode.

The new diode is considerably better from the heat dissipation point of view than is the glass diode because there is little resistance to heat flow

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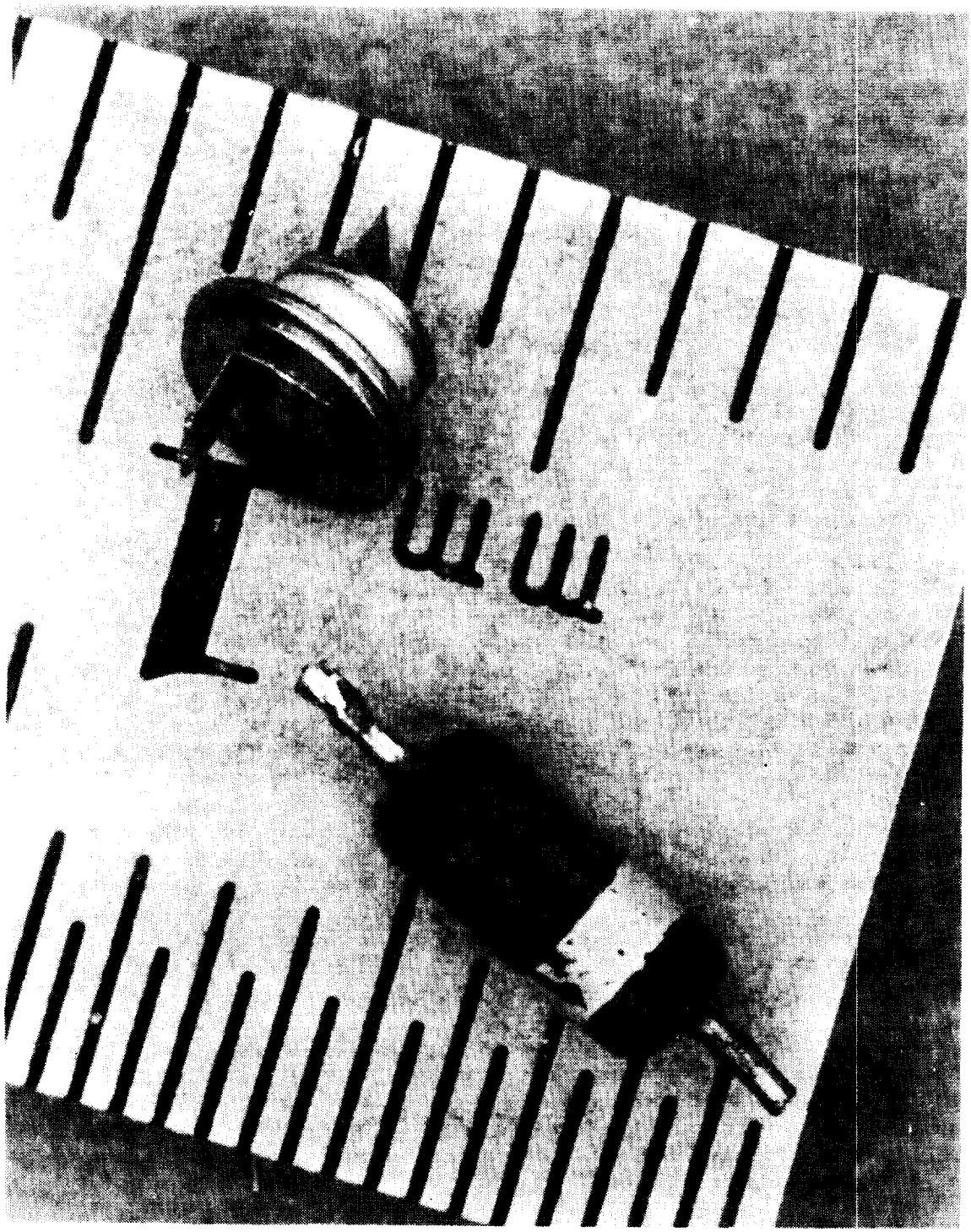

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Figure 2-15. The Diode for the Thin-Film Etched-Circuit Rectenna was Packaged in Two Formats Shown Above. One was the Miniature Glass Package While the Other was the Miniature Ceramic Pill Package. The Glass Package was Used for the Early Development of the Thin-Film Rectenna. The Ceramic Package is Currently being Used.

from the diode to the transmission line upon which the diode depends for the conduction and radiation of heat. In the glass diode there is considerable resistance because of the dumet lead which is a relatively poor heat conductor. It is also of interest that the heat flows out of the ceramic package in both direction in about equal amounts because of the excellent heat conduction properties of the alumina ceramic shell.

The diode specifications that were used for the procurement of the didoes is as follows:

- Semiconductor material - type n GaAs.
- Plated heat sink for good conduction of heat away from junction.
- Metal used at junction - platinum.
- C_{t0} (zero bias capacitance) - 3 pf $\pm$ 10%.
- V_b (reverse voltage breakdown) - 60 - 70 volts.
- Reverse Leakage - 10 microamperes at 80% of breakdown voltage.
- Slope of voltage current characteristic - 1.5 ohms maximum at 100 milliamperes of DC current.
- To be packaged in Raytheon ceramic package #3000119.
- 0.040 inch wide ribbons to be attached to both covers of ceramic package.

The electrical portion of these specifications is considerably relaxed from that used for comparable diodes for the large rectenna array designed and manufactured for a microwave power demonstration on the Mojave desert in 1975. The motivation was reduced cost achieved by eliminating rigorous quality control procedures. However, for any procurement where high reliability is needed it is recommended that a more restrictive specification be used.

In particular, it is essential to specify the voltage drop across the entire diode at very low current conduction in the forward direction, because the slope of the voltage current characteristic at 100 milliamperes does not necessarily guarantee a low voltage drop across the diode. For example, another Schottky barrier can be used in place of a low resistance ohmic contact. While this will automatically double the voltage drop across the diode at low currents and drastically reduce the efficiency of the diode it will not greatly impact the slope of the voltage-current characteristic at high currents.

2.1.7 Summary of Activity on Task 1

- The causes of low efficiency in the thin-film, etched-circuit rectenna element were found and identified as unexpectedly high loss in the mylar and expected losses in adhesive.

- In a search for better film materials Kapton Type F was found and projected to result in a much superior rectenna element in terms of efficiency and ability to operate at high temperatures.

- Three reiterations of the rectenna element design were made before finalizing the printed circuit masks in the third reiteration.

- Substantial changes were made in the Schottky barrier diode design. The package was changed from glass to ceramic; the area of the junction was increased three fold to improve diode yield and to be more in accord with the trend towards operation of rectenna elements at higher power levels for space and high altitude air vehicle applications.

- Electrical tests were made on the final design of rectenna element over a wide range of microwave power input and DC load resistance. Efficiency and reflected power were noted. An overall efficiency of 85% at 1.5 watts DC/output was obtained and a rectification efficiency of 84% was obtained at 3.5 watts of output. Limitation in power output and efficiency was the reverse voltage breakdown of diode.

2.2 Design of a Complete Rectenna Array (Task 2)

2.2.1 Introduction

One of the attractive features of the rectenna for space use is the assumed relative ease with which it may be deployed. For example, in the scenario assumed in Reference 8 all the rectenna that would be necessary for a 50,000 square meter array in space, enough to conservatively generate at least 10,000 kilowatts for an electric powered interorbital vehicle like that shown in Figure 2-16, could be carried into low earth orbit in one shuttle payload.

The rectenna would be carried into space in rolls that are 18 meters long to fit within the shuttle payload bay. Each roll would contain a rectenna that is 200 meters in length wound on a spool that is 30 cm (1 foot) in diameter. The outside diameter of the spool would be 71 cm (2.33 feet). These dimensions allow for a spacing of 0.165 cm (0.065 inch) between turns which accommodates a special construction that allows the rectenna, consisting of a foreplane and reflecting plane, to collapse when it is wound on the spool. Since each of the rolls is 18 meters long, the number of rolls needed to comprise the 50,000 square meter rectenna area is 14. The rolls could therefore be stored within a cross sectional area of 7 square meters or in about 35% of the 20 square meter cross section of the cargo bay. The estimated mass of the complete rectenna is 10,000 kilograms, and a more conservative figure of 15,000 kilograms would require about one half the payload mass of the shuttle.

The foregoing example of deployment in space represents a good starting point for a more general discussion of deployment, and in introducing a general problem in deploying the rectenna. The general problem of deploying the rectenna is that to capture energy efficiently the rectenna must be a two plane structure. Behind the thin-film etched-circuit foreplane there must be a reflecting plane about 0.2 wavelengths or 2.5 cm behind the foreplane. The foregoing example assumes that this distance is collapsed down to 0.165 cm, a ratio of 15.

Alternative means of deployment that come to mind are (1) taking the rectenna up in panel form, (2) deploying the foreplane from one roll and the reflecting plane from another roll, (3) deploying both from the same roll. The first alternative has many objections, including the necessity of very large storage space in the shuttle, increased mass, and high cost. The second alternative resolves the need to collapse the thickness of the rectenna but leaves unresolved the need to maintain the spacing of the reflecting plane from the rectenna foreplane. Both of these options do not seem very attractive.

2.2.2 An Approach to a Collapsible Rectenna

The example of deployment of a collapsed rectenna from a single roll also illustrates that the collapse does not have to be that complete. A complete collapse of the rectenna foreplane and the reflecting plane without the consideration of the diode would be only 3 mils. The eventual diode

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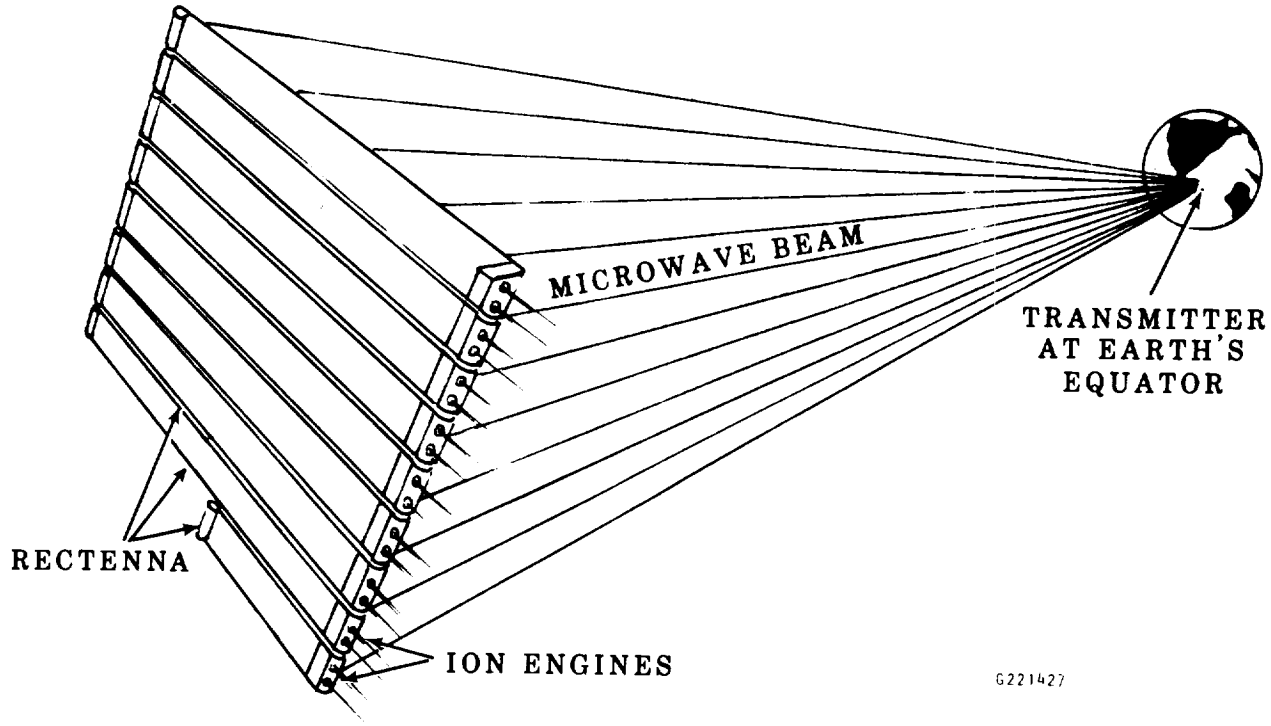

Figure 2-16. Concept of a Microwave-Beam Powered Transportation Mode from Low-Earth to Geosynchronous Orbit. Microwave Power is Beamed from a Ground Station to the Vehicle which has a Large Rectenna to Absorb the Microwave Power and Convert it into DC Power to Energize the Electrical Propulsion Engines. The Vehicle is Transported in Modular Form within the Cargo Bay of the Shuttle and Assembled in Low-Earth Orbit. Each Module Contains Ion Engines Together with Their Supporting Rectenna Sections which are Stowed in Rolled Up Form During Transportation in the Shuttle.

developed for this application should not exceed ten mils in thickness; so that a collapse to 20 mils in thickness should eventually be possible.

However, the present height of the diode is 60 mils. It may be useful to use this as a starting point and assume that the resulting collapsed thickness would be 0.065 inch. This distance of 0.065" then permits about 0.062 inch for the collapse of some kind of compression spring.

In the stowed position, the spring is collapsed but in the deployed condition the spring is still under some compression and its extended length is constrained by a string attached to the foreplane and reflecting plane. The length of the string, fully extended, represents the desired distance between the foreplane and the reflecting plane.

Although the use of a metal spring may be considered, a moment's reflection would suggest that there are potential complications and that a plastic spring might therefore be preferable. Although plastics as a spring material are greatly inferior to metal, the application under discussion does not demand repeating flexing; nor does the spring in its extended position need to exert much force.

In making a search for the best plastic material, it might be well to start with Kapton itself to see how it might be used. Kapton is already being used in the rectenna. It also has some interesting mechanical properties. It has a combination of a yield point and a modulus of elasticity that allows it to be wrapped around a mandrel of small diameter in relationship to the thickness of the Kapton before the Kapton takes a permanent set. Or, expressed another way, the ratio of the radius of curveature to the thickness of the material before it takes a permanent set is relatively low. Low enough to suggest that rolled up circular loops of Kapton film two to six mils thick may be flattened to 60 mils; and then spring out to their original diameter, or near it, when the compression is released.

An experimental program to investigate this was undertaken. Kapton of different thicknesses was obtained, rolled up into loops, the ends of the loops sealed together with scotch tape. These loops were then compressed between two flat metal plates. One of the plates had two side rails of 0.062" thick material to limit the amount of loop compression to that assumed that it would undergo when being wrapped up in a roll. With this amount of compression it was found that three mil Kapton took no set at all, while 5 mil Kapton took a small amount of set.

After the loops were so compressed, they were mounted between two plates, the top plate being very light in weight and free to move with respect to the bottom plate. Weights were added to the top plate and force-deflection data taken on the loops. From these force-deflection curves, it was noted at what force the height of the loop was 2.5 cm, the expected distance between the

foreplane and reflecting plane of the rectenna. Pepresentative data obtained from this experimental procedure is given in the following table.

Compression characteristics of circular loops of Kapton.

Diameter of Loop (cm)	Thickness of Kapton (mils)	Force Required to Compress Loop to the Normal Separation Distance of 2.5 cm Between Foreplane and Reflecting Plane (Grams per cm Loops Width)
4.13 (1.625 inch)	5	6.2 grams
4.13 (1.625 inch)	3	1.0 grams
5.08 (2.0 inch)	5	5.3 grams

From this data it is noted that a loop one centimeter wide made from 5 mil Kapton exerts a force of five or six grams between the two planes at a separation distance of 2.5 cm. In a "gravity free" space environment, it is assumed that this is enough force to always keep the constraining strings taut. How many of these would be needed per unit area would have to be determined eventually by some experimental work performed in a gravity free environment.

Figure 2-17 shows how such loops of Kapton could be applied to the foreplane of the rectenna. Each of the loops would have a length of string inside it that corresponded to the desired separation distance between foreplane and the reflecting plane.

2.2.3 Considerations Involved in Spooling the Collapsed Rectenna

From the past discussion it appears that there is a straight forward method of maintaining the separation distance between the reflecting plane and the foreplane that still permits compressing the rectenna to a thickness of 0.165 cm (0.065 inch). However, it must be recognized that when the rectenna is rolled up there will be a difference in the circumference of the two surfaces equal to 6.28 times the separation distance between the two planes. This distance is accumulative over the entire length and could amount to several meters.

There are two potential solutions to this problem. One solution is to roll the reflecting and foreplanes together but leave the reflecting plane unattached. In this case some force must be found to press the reflecting plane against the supports extending from the foreplane. One mechanically simple way is to use electrostatic attraction between the two films that could be established by a voltage applied between the two surfaces. A potential of 1000 volts would provide an attractive force of several grams per square meter, enough to hold the reflecting plane to the foreplane supports.

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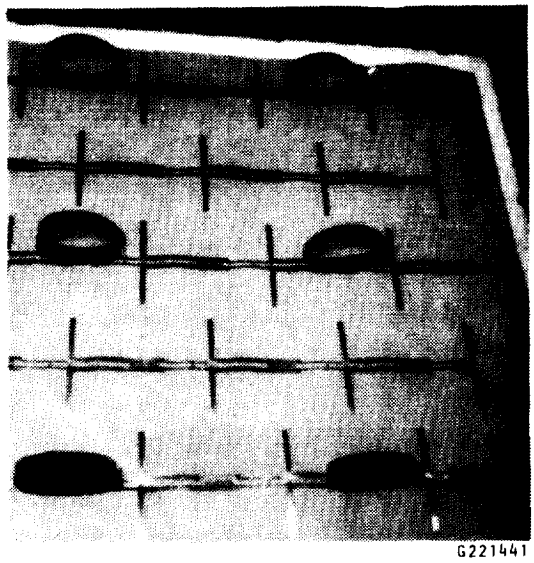

Figure 2-17. Photograph of Rectenna Foreplane with Loops of Kapton Attached to Separate Foreplane from Reflecting Plane. Loops are Compressed During Roll-Up on Spool, but Return to Original Shape when Compression is Released. A string in Center of Each Loop Limits Height to Separation Distance Needed between Foreplane and Reflecting Plane.

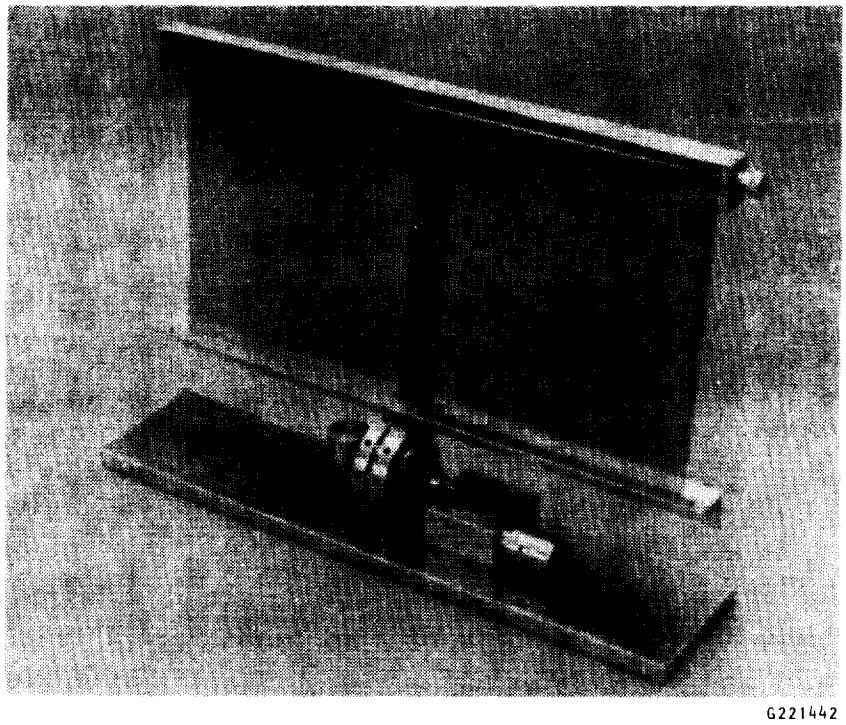

Figure 2-18. Model Showing Deployment of "Rectenna" from Rolled Up to Extended Form by Means of a Support Member which can also be Rolled Up but which becomes Rigid when Unrolled. Model Used a Carpenter's Rule to Illustrate the Principle.

A second approach would be to attach the planes to each other but to wind up the rectenna while retaining some slack in the reflecting plane in the rolled up condition. Still another alternative is to put the reflecting plane toward the outside and deliberately apply some tension to it while it is being rolled up. The reflecting film is the film that should be manipulated because it will have a small fraction of the mass and strength of the foreplane.

Related Considerations Involving Support of Rectenna in Space When Unrolled

The selection of a construction format for the two plane rectenna depends not only upon how it is stowed before deployment but also upon how it is deployed. In general, there would be two kinds of deployment. The first would be to unroll it onto some kind of a supporting structure. The second would be to have the rectenna combined with some mechanical boom that would also roll up into a cylinder immediately adjacent to the rectenna. At this time it is too early to determine which is the preferred method, and the method used may well depend upon the space application.

The second mode of deployment is that portrayed in Figure 2-16 and in a photograph of a model of such deployment (Figure 2-18). The model uses a thin, curved, metal strip that is rigid when extended but that can be rolled up to conserve space. It is, in fact, a commercially available 6 foot carpenter's rule. In space, of course, with no steady state forces acting on the boom, such a boom might be 100 meters or more in length. It should be possible by means of tethers to slightly bow the boom so that the reflecting plane would be on a convex surface. In that case, the reflecting plane could be kept in contact with the separators by a slight tensile force exerted at the two ends of the reflecting plane.

The first mode of deployment might very well be used in connection with a much larger structure that would either be a permanent low earth orbit around the equator, or that could be used as a heavy transport vehicle for orbital transfer. Such a structure if it were used for manufacturing might receive all of the electrical power it needed from transmitters located at the equator on the surface of the earth. Or if used for interorbital transport, such a large aperture would make it efficient and useful well beyond geosynchronous orbit, perhaps even to the moon. Such a platform would no doubt have some kind of rigid support truss. This would make for easy deployment of the rectenna and forming it into a slightly convex surface to permit stretching of the reflecting plane over the rectenna.

2.3 Fabrication and Testing of a Large Area Sample Rectenna (Task 3)

2.3.1 Introduction

In several respects, the fabrication and testing of a large area sample rectenna was the principle objective and task of the first activity covered in this final report. The objective is to assemble rectenna elements of an identical optimized design, into a very light weight (low mass) rectenna foreplane of considerable area that can be easily fabricated and that will operate at high efficiency and at relatively high power density levels with long life and high reliability.

Although a 14 element foreplane had been fabricated from a design based on the use of mylar as the thin film material under the Wallops Flight Center study only individual elements had been tested, and these had been found to be inefficient, as reported in Section 2.1.

During this new study we were able to first fabricate and test a 14 element foreplane made from a much improved sandwich material using Kapton-F film and employing diodes having improved power handling capability. No measurements of diode temperature or velocity of cooling air were made in these tests.

Then new masks were made for a larger rectenna area consisting of six rows of five elements each, or a total of 30 elements. For test purposes only, five rows of rectenna elements were used as shown in Figure 2-19.

The test set up for this 25 element rectenna was instrumented to make measurements of the velocity of the cooling air flowing over the surface and of the operating temperature of the diode case of the central rectenna element. A DC output of 120 watts from this 25 element section was achieved with a temperature rise of only 50°C with an air flow velocity of 4.8 m/sec., 15.8 ft./sec., or 10.8 Mi/hr.

The setup for testing the 25 element section was then used to test a single rectenna element. First, without microwaves being involved, measurements were made of the temperature rise of the diode as a joint function of (1) DC power injected into and dissipated within the diode and of (2) the velocity of the cooling air flowing over the rectenna. Then, with a known velocity of air flowing over the element, the microwave power beam was directed at the rectenna element and measurements of DC power output and diode temperature noted. From the use of tabulated data in the previous run which related diode dissipation to diode temperature, the rectification efficiency of the rectenna element could be approximated closely from the expression

$$\text{rectification efficiency} = \frac{\text{DC power output}}{\text{DC power output plus diode dissipation}} \quad (1)$$

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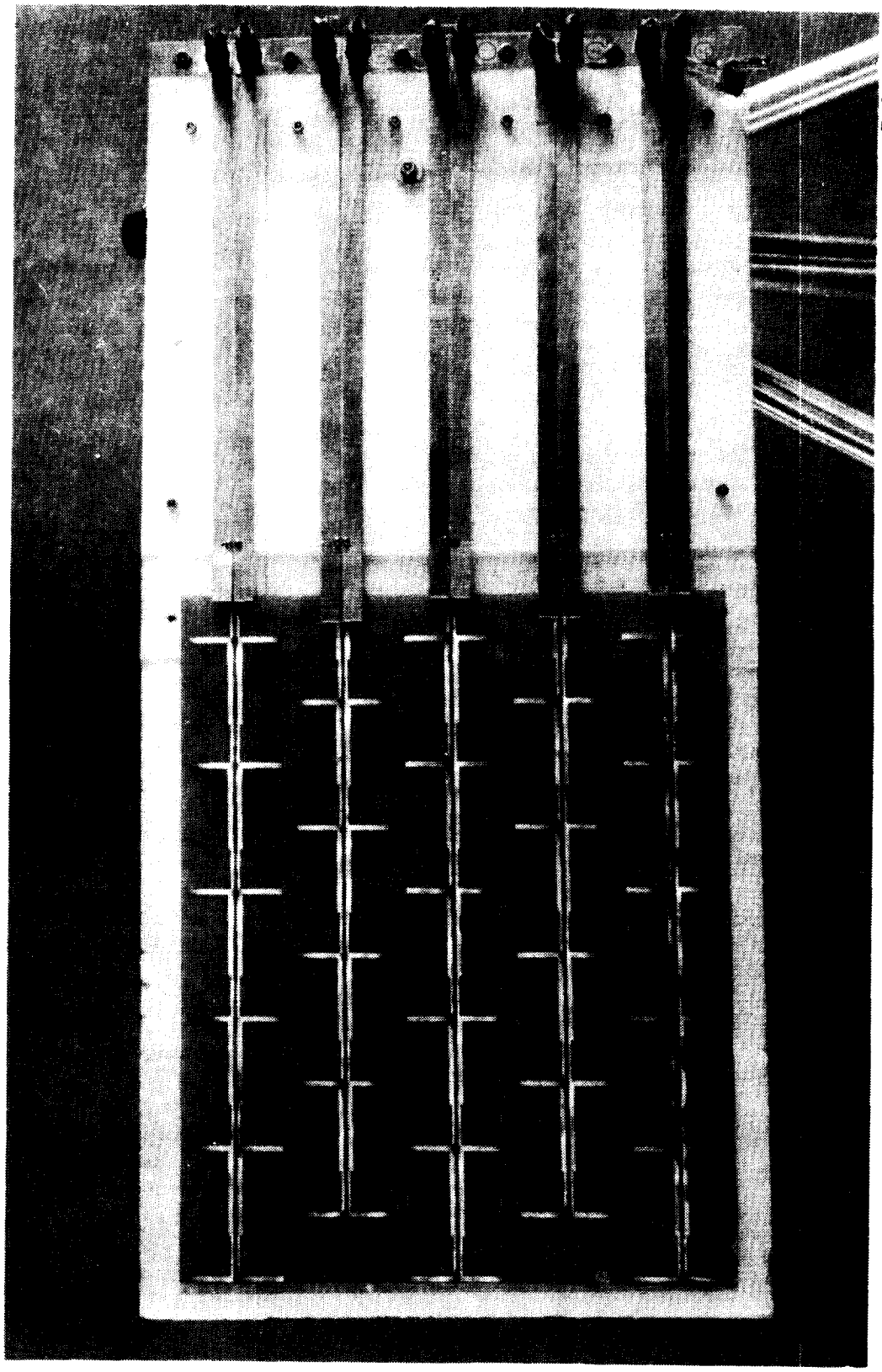

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Figure 2-19. Rectenna Array with 25 Active Elements Set Up for Test in Microwave Beam as Shown in Figure 2-20. Foreplane is Separated from Metallic Reflecting Plane by 1.8 cm of Styrofoam. Three of the Five Inboard Elements on the Rectenna are Not Active.

These efficiencies were consistently over 80%. Power output as high as 6 watts for the element was also obtained. Maximum power output was limited not by the diode temperature but by the value of reverse voltage breakdown of the diode and by the high characteristic impedance of the rectenna element structure. With a proper redesign of the diode and the balance of the rectenna element, twenty watts per rectenna element could be achieved without exceeding either the allowable operating temperature of the diode or the velocity of air flow used in these tests.

In defense of this procedure for measuring the rectification efficiency of a single rectenna element, it is noted that the edge effects are not well known and would introduce a large uncertainty into an efficiency calculated on the basis of measuring the incident power density of the beam.

2.3.2 Test Arrangement and Test Results on 25 Element Rectenna Foreplane with Forced Convective Cooling

This section discusses test results on power handling capability and efficiency of the rectenna foreplane when subjected to low-velocity convective air cooling at sea level air densities. These conditions are equivalent to cooling obtained at typical airplane flight speeds and air densities at altitudes of 40,000 feet and more.

The tests were made with the test configuration shown in Figure 2-20. The velocity of the air stream flowing across the rectenna was calibrated with a hot wire anemometer. The temperature of the diode was monitored during operation with a probe of a fluoroptic thermometer that was mounted at the diode case with precaution to shield the probe from the air stream. The great advantage of the fluoroptic thermometer is that it does not interact with the microwave field and does not conduct heat away from the source it is measuring.

In preparation for testing, the 25 element rectenna section was mounted on a test board as shown in Figure 2-19. The DC outputs of the five rows of rectenna elements were connected in parallel across a single load resistor. Each row had a zener diode shunted across it to protect the diodes in the row should the DC output voltage for any reason exceed a value that would cause the peak inverse voltage across the diodes to exceed their reverse breakdown voltage rating. An open circuit concurrent with large values of incident microwave power would be the type of fault for which the zener diodes would provide protection.

A current jack in the output of each row made it possible to monitor the operating current and power output of each row.

The array shown in Figures 2-19 was then inserted into the test arrangement depicted in Figure 2-20. So as not to exceed the reverse voltage breakdown of the diodes at high incident power, the common load resistance used was 4.7 ohms. If the 4.7 ohm resistance was equally divided among the 25

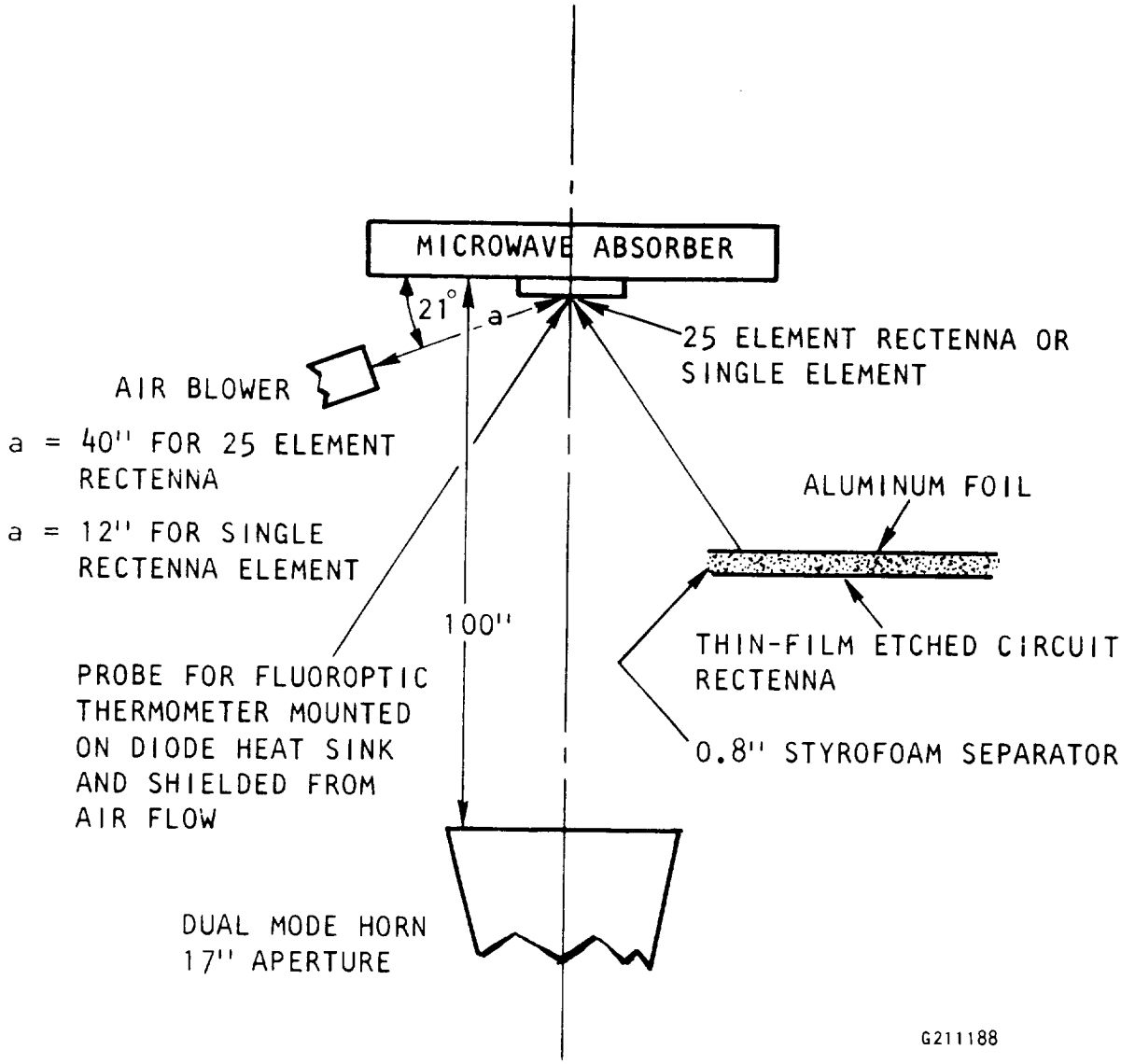


Figure 2-20. Test Arrangement for Measuring DC Power Output and Diode Temperature Rise of 25 Element or Single Element Rectenna as Function of Laminar Air Flow Velocity and Incident Microwave Power Level.

elements, each element would see a value of 118 ohms. The air flow velocity was 4.8 meters/second. The temperature of the case of the diode of the central rectenna element was monitored. For the test, the microwave power output of the illuminating dual-mode horn was varied and the DC power output of the rectenna and the temperature of the diode case were monitored.

The results of this test are shown in Figure 2-21. The maximum total power output of 123 watts corresponds to an average of 4.9 watts for each of the individual diodes. The unequal illumination of the diodes as well as the tendency of the top and bottom rows to capture more power than the internal rows means that some of the rectenna elements were producing more than 6 watts of DC power output. The highest diode temperature of 88°C is far below the 200°C operating temperature considered safe for GaAs diode operation. The temperature rise in the diode was approximately 48°C. The high temperature of 39.5°C for the diode with no incident microwave power was caused by heat stored in the microwave absorber that was placed behind the rectenna and that had absorbed energy over a period of time prior to making the run recorded in Figure 2-21. The room ambient temperature in regions removed from the absorber was 31°C.

The observation of such high DC power outputs with only relatively low air flow velocity and with such low temperature rise in the diode was an exciting experimental finding that had not been expected. It is an especially important finding for many of the applications of microwave power transmission in both aircraft and space.

The normal procedure, with the data of Figure 2-21 in hand, would be to operate the rectenna at even higher power densities, but unfortunately, the reverse voltage breakdown of the diode and the relatively high impedance design of the rectenna element precluded higher power operation. However, as indicated in the introduction it is possible to extend the data for temperature rise in the diode as a function of power dissipation within the diode and air flow velocity to provide an accurate indication of the DC power output capability of the rectenna element, if an efficiency for the element is also assumed, as will be discussed in Section 2.1.3.

A finding of considerable importance was made in testing the rectenna. It was found necessary to put all the rows in parallel. If they were put in series there was an unstable load sharing situation in which some rows capture a large amount of power while others capture very little and put out very little DC power. When the rows are in parallel there is good load sharing particularly at the higher power levels where presumably each individual element is better matched into space.

It would also be expected that the two rows of elements at the edge of the array would pick up more power than the center rows and this was found to

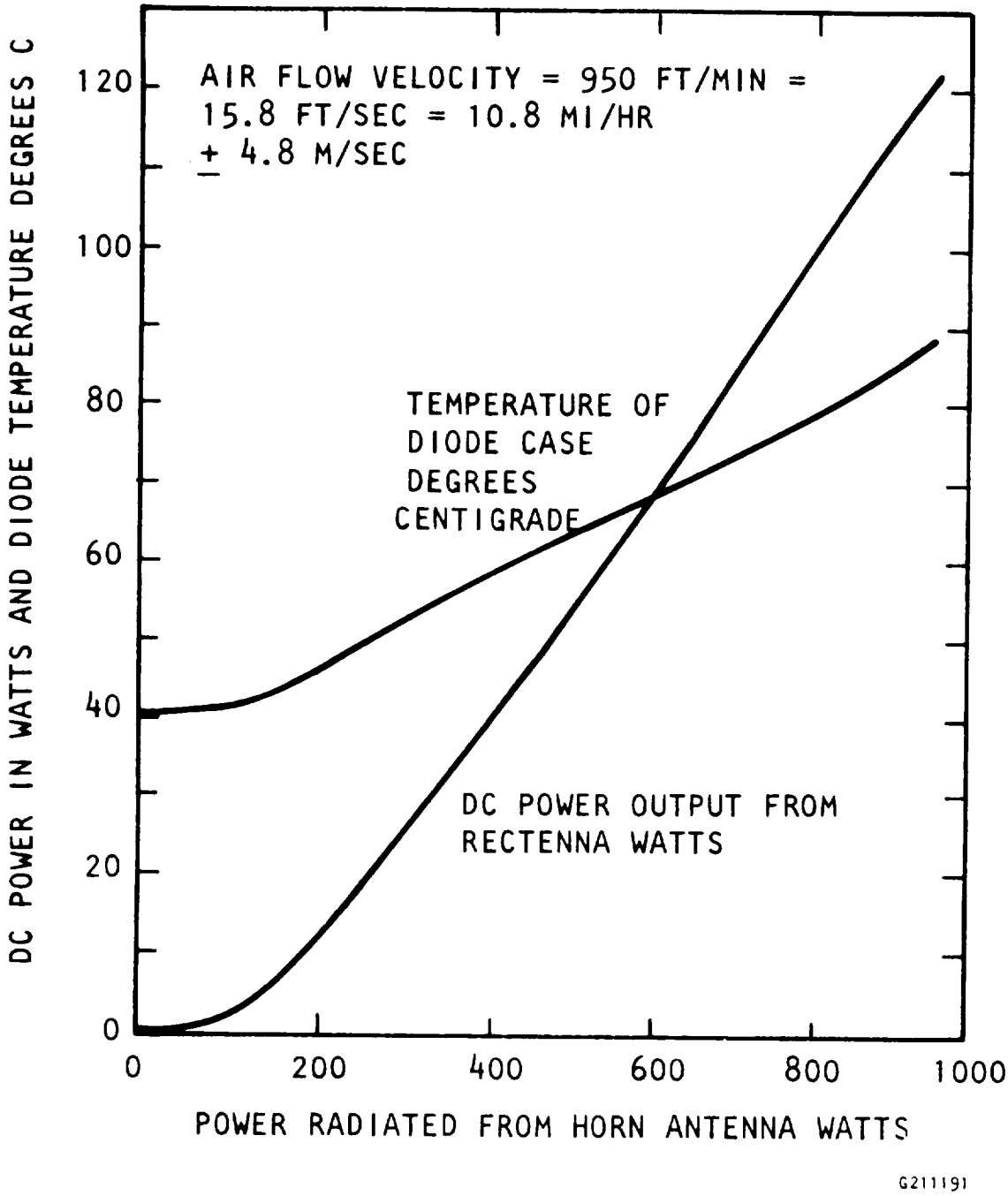


Figure 2-21. DC Power Output of 25 Element Rectenna and Case Temperature of Diode of Center Element as Function of Total Power Radiated. Only a Small Fraction of Radiated Microwave Power Impinged Upon the Rectenna. Rectenna Efficiency cannot be Computed from Data Taken.

be the case experimentally as given by the following data, where Row 1 and Row 5 are the outside rows.

	Row 1	Row 2	Row 3	Row 4	Row 5
DC Power Output	7.4	5.5	5.9	5.8	7.8

These ratios of power levels of outside strings to inside strings was found to be about the same for total DC power outputs ranging from 30 to 120 watts.

In addition to the non-uniformity of pick up of outside and inside rows, the microwave beam itself was not uniform. An attempt was made to illuminate the 25 element rectenna uniformly using a sufficient separation from the mouth of the dual mode horn to allow the gaussian beam emitted by the horn to expand to a diameter so that the intensity of the radiation across the 25 element rectenna would be relatively uniform. On the other hand, the separation distance cannot be too great because the power density of the incident beam on the rectenna could fall below that needed to evaluate the rectenna at high power level. The distance of 100 inches proved to be a good compromise. The power density at the corners of the rectenna were 0.8 below that at the center, while the maximum illumination used on the rectenna necessitated nearly 1000 watts of power being radiated from the dual mode horn.

2.3.3 Temperature Rise of Diode in Rectenna Element as Function of Injected DC Power and Velocity of Convection Cooling Air

The injection of measured DC power into the diode and dissipated there provides a heat source of known value. This heat is generated in the same area as would be heat from microwave rectification losses, and it is conducted in the same manner as would be heat generated by microwave rectification into the same radiatively and convectively cooled structure. Because the dissipation in the diode represents 80% of the total heat generated in the rectenna element, when it is acting to convert microwave energy into DC power, these other sources of heat generation were ignored in computing efficiency. It is noted, moreover, that these other losses are much more uniformly distributed throughout the rectenna element and tend to dispose of the associated heat more efficiently than the diode. Their impact upon temperature measured at the diode is therefore limited.

The experimental arrangement for taking the data is shown in Figure 2-20. The data that resulted from noting temperature rise in the diode as a function of injected DC power and the velocity of the convective cooling air stream flowing by the single rectenna element is shown in Figure 2-22. The linearity of the data given in Figure 2-22 is an indication that the air flow over the rectenna surface remained laminar for all of the test conditions.

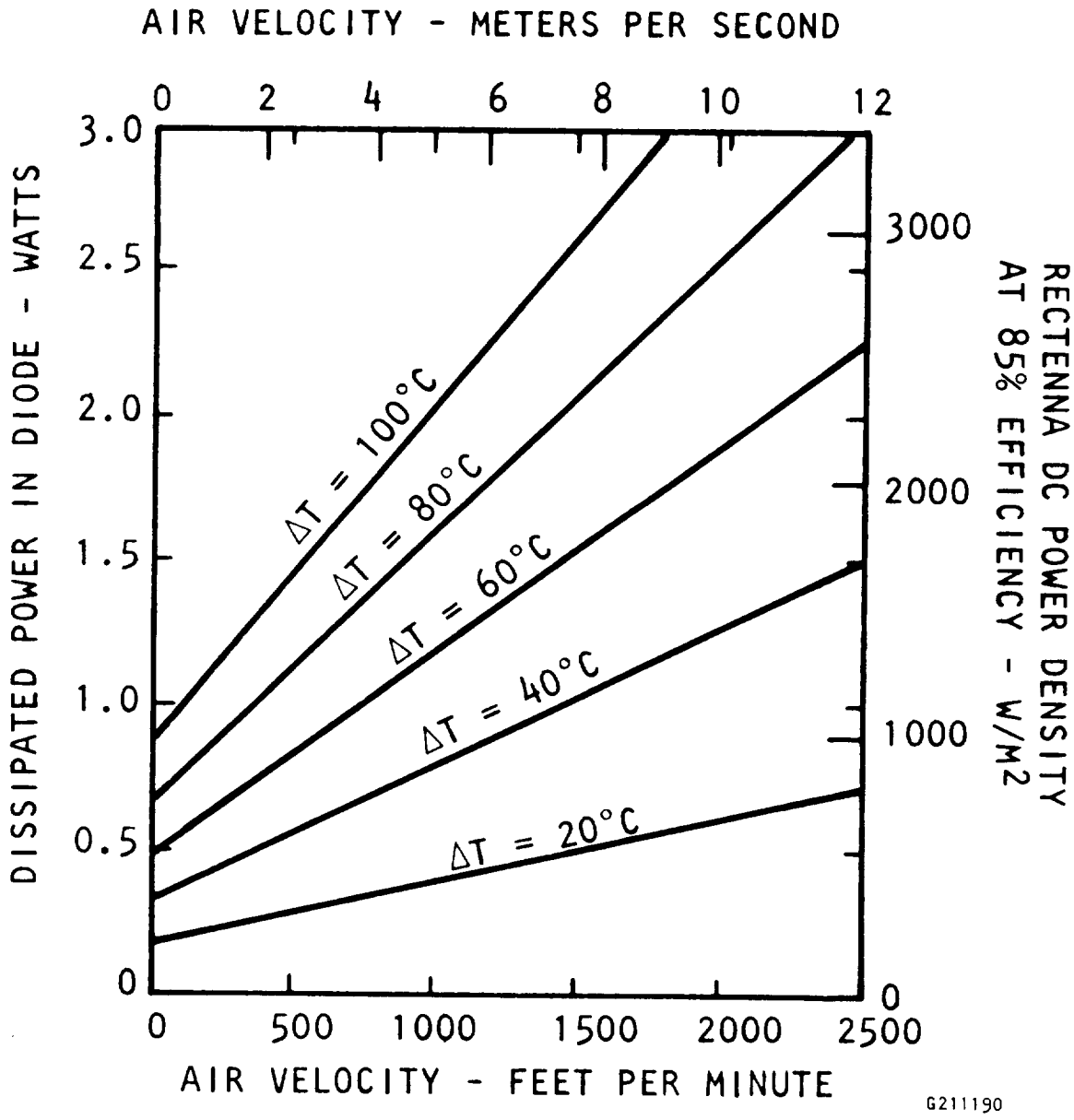


Figure 2-22. (a) Power Dissipated In Single Diode, as Function of Air Velocity and Diode Temperature (Left Hand Scale). (b) Equivalent Rectenna DC Power Density at 85% Efficiency (Right Hand Scale).

The data in Figure 2-22 can be used to develop equation 2 for diode dissipation as a function of air velocity and temperature rise of the diode case. The equation has two components. One component is that associated with convective cooling. The other is associated with residual cooling at zero air velocity and is assumed to be largely radiative cooling. Note that the linearity of the data and the low Reynolds number associated with the air velocity indicates laminar air flow at all time.

$$W = 1.125 \times 10^{-5} \Delta T_1 V + 0.009 \Delta T_2, \text{ where} \qquad (2)$$

W = power dissipated in diode in watts

V = air velocity in feet per minute

ΔT_1 = °C increase in temperature of diode case above ambient air temperature

ΔT_2 = °C increase in temperature of diode case above temperature of sink for radiated power

Now if we know or make an estimate of the rectenna element efficiency, n, we can multiply equation (2) or the ordinate data on the left side of Figure 2-22 by the factor n/1-n to obtain the DC power output of each rectenna element. For an assumed efficiency of 85%, a typical value for the rectenna element operated in the one to five watt output region, this factor will be 5.7. The corresponding DC power for one square meter which contains 200 elements is shown as the right hand ordinate of Figure 2-22 as a function of air velocity and diode temperature rise.

Equation (2) was derived from experimental data using mass flow rates of air at sea level density. If the term V in equation 2 is replaced by $V\rho/\rho_o$ where V and ρ are the air velocity and air density at any altitude and ρ_o is the air density at sea level the same mass flow rate of air is maintained. If the assumption is made that convective cooling is proportional to mass flow rate of air alone, then the modified expression is applicable to any altitude and can therefore be applied to high altitude airplanes or balloons.

The expression becomes interesting when applied to very high altitude airplanes with the rectenna exposed to the air flow. Although the air density at an altitude of 20 kilometers is only 5% that at sea level, an airplane must fly at a velocity of about 70 meters/sec. to maintain altitude. This would be equivalent to an air velocity of 3.5 meters/sec. (714 ft./min.) at sea level. Figure 2-22 indicates that at this sea level velocity, and an assumed operating temperature of the diode case of 100°C above ambient, nearly 2000 watts of DC power output could be obtained from the rectenna, if the additional assumption is made that the diode efficiency is 85%.

The assumption of diode efficiency of 85%, corresponding to an inefficiency of 15%, is a conservative one. In Section 2.0 of Reference 6, diode losses were accurately measured to be 8% of the incident microwave power when the diodes were operating in the 6 to 10 watt region.

Although a minimization of cost analysis of microwave power transmission systems for the applications examined to date favor a moderate DC power density level from the rectenna, it should be noted that the power output from a single rectenna element could, with convective cooling, be easily upgraded to 25 to 50 watts of DC power output. This could be achieved by redesigning the rectenna element to match into a much lower DC load resistance, of the order of 30 ohms. The redesign should be relatively straightforward and would result in an optimum diode that would have a much larger diode junction area further increasing the power handling capability of the diode.

The final test performed on the single rectenna element was to illuminate it with microwave energy and note the DC power output and the temperature rise in the diode at a fixed air flow velocity. From this data, and with the use of Figure 2-22 and equation (1), it was possible to compute the operating efficiency of the rectenna element. The resulting data is shown in Figure 2-23. The similarity of this data with that obtained from a closed system for the rectenna element and presented in Figure 2-14 is of interest.

The technique of measuring diode temperature as a function of DC input power dissipated within the diode is of particular significance in determining the power handling capability of the rectenna in vacuum without having to apply any microwave power in vacuum. The procedure would be to first check the efficiency of the rectenna element outside the vacuum using test equipment similar to that shown in Figures 2-2 and 2-3.

The rectenna element is then inserted into the vacuum environment and its case temperature noted as a function of DC power injected into the diode to be dissipated there. Then at a specific case temperature, the power dissipation is known and the DC power output under assumed microwave radiation conditions can be found from the expression.

$$\text{DC Power Output} = \frac{n}{1 - n} \text{ (Power Dissipated in Diode)}$$

where n is the measured efficiency outside the vacuum.

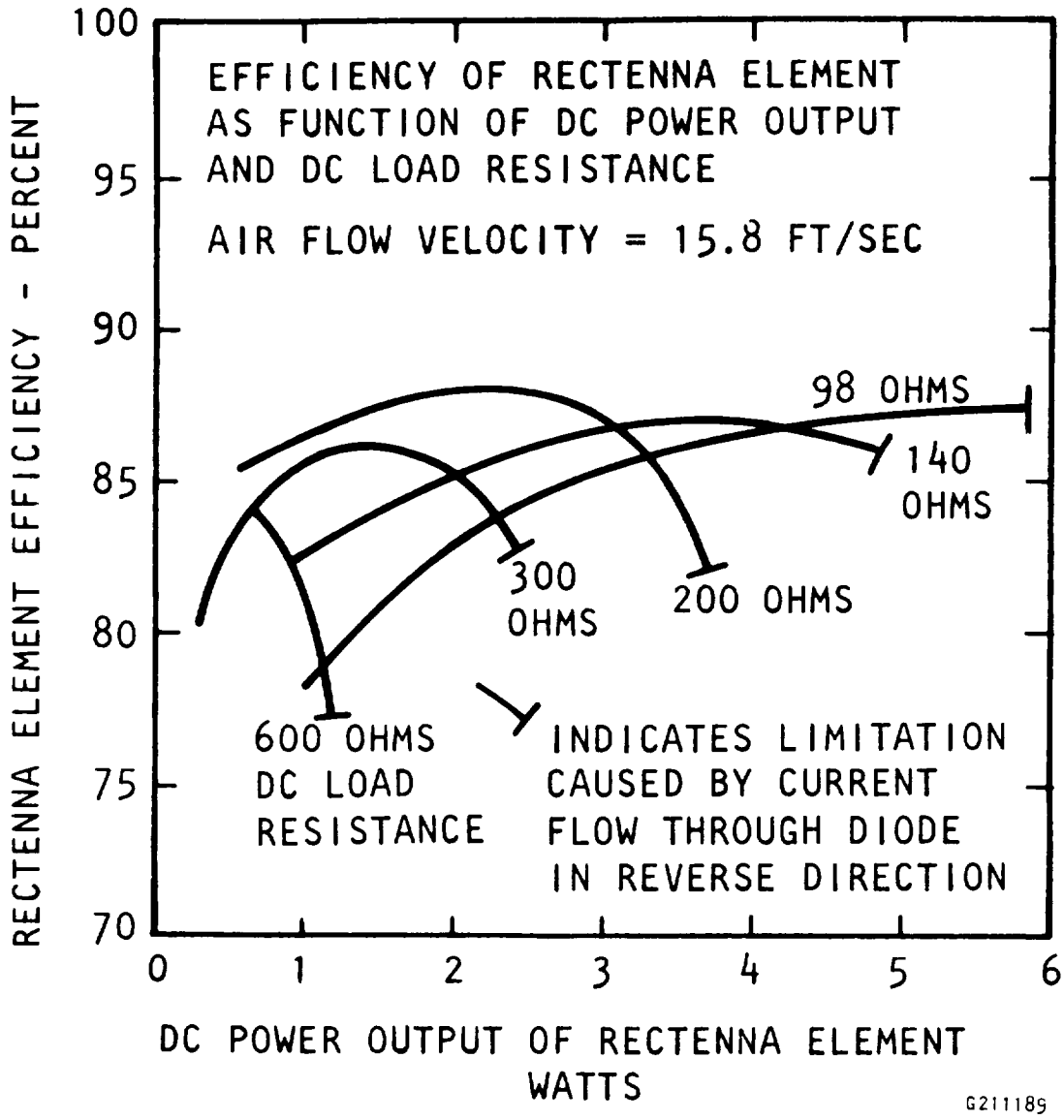


Figure 2-23. Efficiency of Rectenna Element as Function of DC Power Output and DC Load Resistance as Computed from Measured Dissipation in Diode and DC Power Output. Compare with Data of Figure 2-14 for Same Rectenna Element Measured in Closed Test Fixture in which Microwave Power Input could be Accurately Measured.

2.4 Design of an RF Test Facility (Task 4)

This task concerned the design of a suitable rf test facility to be built at Lewis Research Center for the purpose of testing the rectenna. The task was concerned not only with advice about the rf test equipment to be used but also with certain aspects of the testing procedure. A major consideration was that the tests were to be made in vacuum.

A testing procedure to be carried out in vacuum requires the measurements of DC power output from the rectenna and the operating temperature of the rectenna, with special emphasis upon noting the operating temperature close to the rectifying diode which will be the major source of heat and which is the most susceptable to heat damage.

It is also desirable to know the operating efficiency of the rectenna but it may be difficult to measure the incident energy of the microwave beam inside the vacuum.

However, it is possible to obtain efficiency values indirectly but quite accurately by passing a known amount of DC power through the diodes from an external source and noting the temperature rise of the diode case as noted in Section 2.3.2. With such a calibration, the efficiency of the rectenna with microwave power incident upon it can be calculated by noting the DC power output and the dissipated power that corresponds to the temperature at which the rectenna is operating.

The efficiency is given by:

$$\text{Rectenna Efficiency} = \text{DC Power Output}/(\text{DC Power Output} + \text{Dissipated Power})$$

A similar procedure of passing DC current through the diode for calibrating losses in the diode was used in the execution of a previous contract with LeRC.[6]

Considerable thought went into the matter of a suitable source of microwave power. The objective was to provide an economical source of power that can range from only a few milliwatts of power to as much as several hundred watts. The latter may be needed for subsequent testing of relative large sections of rectenna within the vacuum. The only economical way that such a goal can be approached is to make use of the microwave oven magnetron, which is readily available at low cost. A block diagram of the equipment involved in using the magnetron is shown in Figure 2-24.

The approach to achieving the immediately needed low power portion of the wide range power was to operate the magnetron into a matched waveguide resistive load through a ferrite circulator and to tap off a small portion of the total power through a probe penetrating into the waveguide. The power

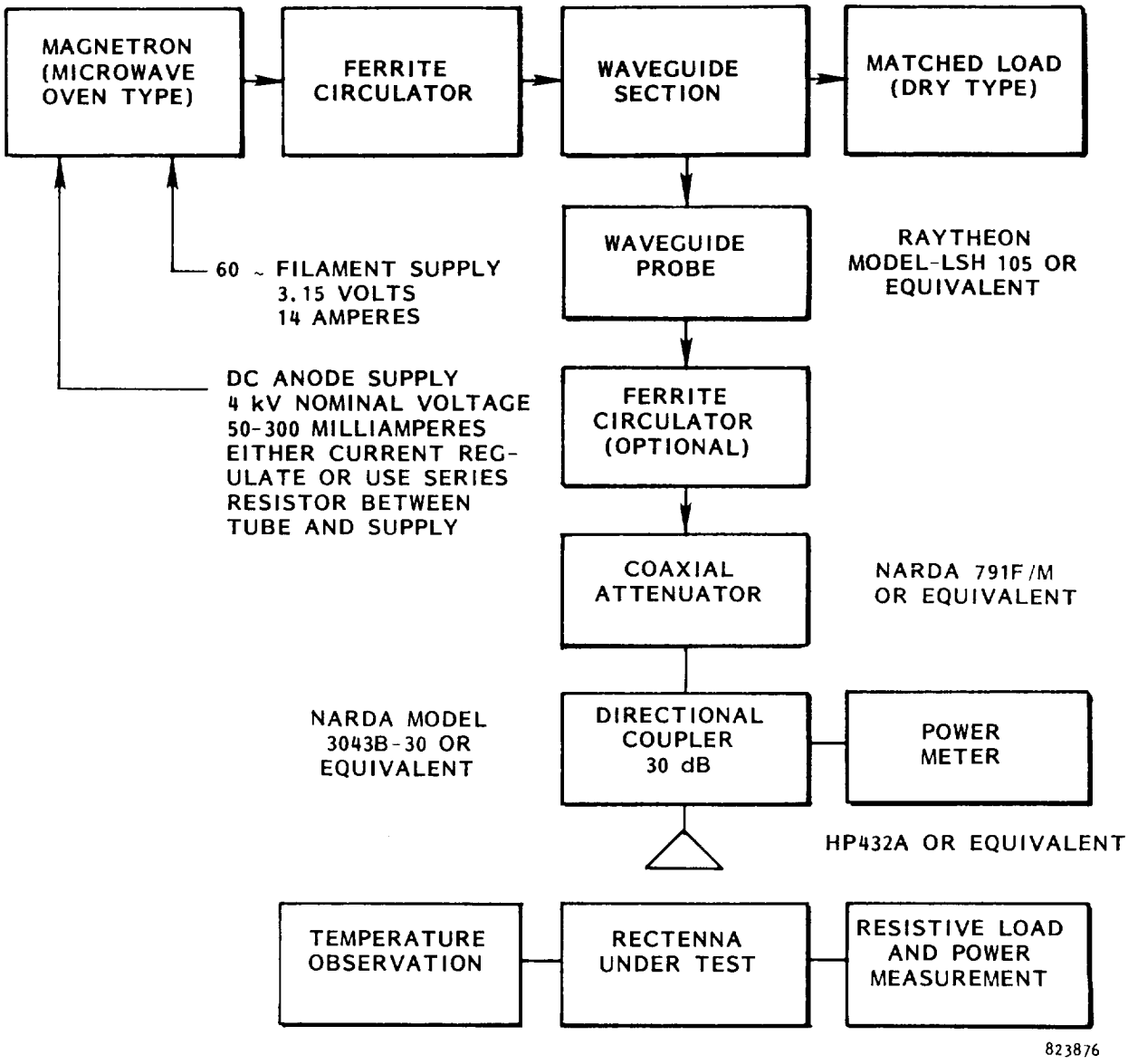


Figure 2-24. Schematic of Test Arrangement.

picked up by the probe could be varied by the length of the probe and by the power level at which the magnetron was operated.

Once the power is picked up by the probe terminating a coaxial fitting, the power can be readily measured in a coaxial system that makes use of coaxial directional couplers and attenuators. This equipment is readily available in most laboratories, but it is also available at reasonable cost from many suppliers. The desired power may then be fed into a gain horn, open ended waveguide, or some other means of funnelling the power into the vacuum chamber. Substantial power may be reflected from the window, depending upon its thickness and dielectric constant. This power may be re-reflected from the probe end of the coaxial system and upset the power reading of the directional coupler. It would therefore be desirable to have a small coaxial ferrite circulator inserted between the coaxial probe fitting and the rest of the coaxial system.

The coaxial system can be used up to a power level of at least 25 watts, which should be adequate for the planned initial testing of a single rectenna element. For higher power levels the waveguide resistive load may be removed and the power fed directly into a suitable radiating device. The waveguide circulator will then absorb any power that may be reflected. If the reflected power level is high it will be necessary to apply a small fan to the resistive elements of the ferrite circulator.

In addition to helping plan the testing arrangements at LeRC, a magnetron, waveguide, and ferrite circulator were sent to LeRC without charge and with recommendations to buy a DC power supply for the magnetron. However, LeRC elected to a buy a PGM-10X. This piece of equipment contains a DC power supply and a 90 watt CW magnetron. This power level is adequate for a broad range of testing, but for higher levels of testing the 500 to 1000 watt output of a microwave oven magnetron may be desirable.

2.6 20 GHz Printed Circuit Rectenna Study (Task 6)

2.6.1 Introduction

The purpose of this study was to investigate different approaches to constructing a rectenna at 20 GHz, and to recommend a specific approach for which the design of a single rectenna element would be developed under Task 7 (Section 2.7 in this report). Two approaches or formats were considered in the proposal submitted. One of these was to scale the present 2.45 GHz thin-film etched-circuit format to 20 GHz. The other was to use an interdigital format to form the capacitances and to keep all circuits on one side of the film to avoid a difficult dilemma in directly scaling the present structure. Both approaches would require the use of a beam lead diode.

After starting the actual study, however, it became quite clear that other approaches should be considered. Among these was a completely monolithic approach in which the diodes as well as all the circuitry are built up on a substrate of gallium arsenide or silicon. This would certainly be necessary at very high frequencies but the technique may be excessively expensive in the 20 GHz region.

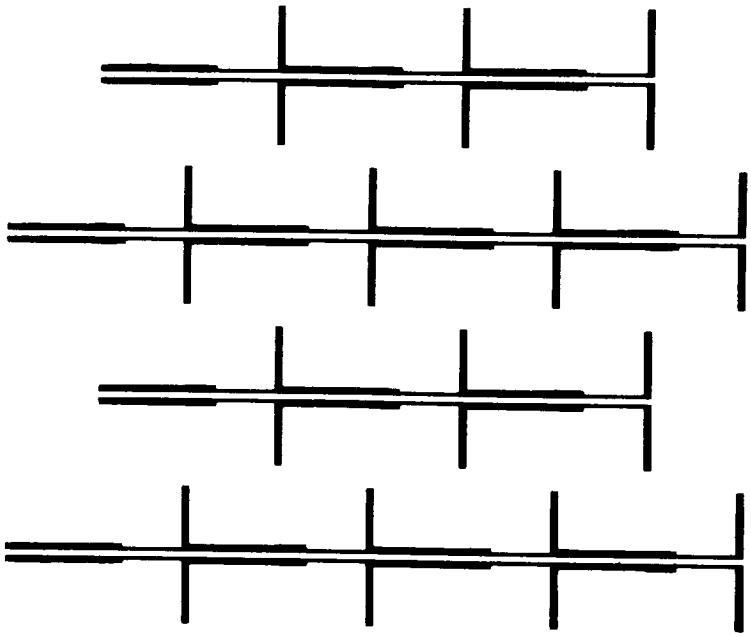
It appeared that a good compromise might be to build a 20 GHz structure on an alumina ceramic substrate which closely resembles a semiconductor substrate in terms of dielectric constant. Beam lead diodes would then be bonded to the resulting structure. The approach was additionally attractive because of current thin and thick film technology based upon the use of alumina ceramic.

These different approaches are discussed in more detail in the following material, but it may be helpful to first approach the study from a more general point of view which is done in Sections 2.6.2 and 2.6.3. Section 2.6.2 discusses the frequency scale in terms of geometry, packing density, etc. Section 2.6.3 discusses the impact of any planning for future higher frequency scaling.

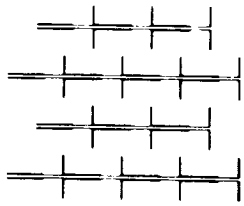
2.6.2 Significance of the Frequency Scale

The starting point for this study is the rectenna technology that exists at 2.45 GHz. However, the shift in frequency, a factor of 8.16, is so great that parameters of the resulting structure at 20 GHz are greatly modified.

Concentration of dipole antennas and diode elements. An exact scale of the 2.45 GHz structure to 20 GHz results in 13,300 elements per square meter instead of 200. The scale at 20 GHz along with a more reasonable scale at 8 GHz is shown in Figure 2-25. If placed on a high-dielectric material, the concentration of elements may be considerably greater than that shown in Figure 2-25 as examined in Section 2.7.3 and 2.7.4, perhaps as big as 30,000 elements per square meter.



2.45 GHz Rectenna Scaled to 8 GHz



2.45 GHz Rectenna Scaled to 20.0 GHz

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Figure 2-25. Large Scale Rectennas can be Scaled to Shorter Wavelengths by Simple Photographic Reduction. Above Material Illustrates Approximate Size of Rectenna for Different Frequencies when Scaled from Existing 2.45 GHz Format.

Spacing between foreplane and groundplane. An exact scale from the present 2 cm spacing at 2.45 GHz would place the new spacing at 0.245 cm or 96 mils. If a high-dielectric material were used to separate the foreplane from the reflecting plane, the separation will be even less.

Dissipation density capability. The power dissipation density increase because of the greatly increased number of rectenna elements even though the dissipation capability of the individual rectenna element is less.

If cooling the foreplane structure is aided by thermal conduction coupling to a continuous reflecting plane, the dissipation density can be greatly increased over that of the current 2.45 GHz thin film structure.

Efficiency. Efficiency must necessarily decrease. The losses in the diode will be considerably greater and the losses in the scaled circuit structure will become greater by the square root of the frequency scaling factor because of skin losses. The importance of efficiency in reducing heat dissipation problems is offset to some degree by the more effective conduction cooling of the scaled structure.

Power density levels. It may be that to keep the diode efficiency high it will be necessary to operate at DC power output levels per element of over 100 milliwatts. But the scale factor then would place the minimum DC power output at a power density level of 1340 watts per square meter. The DC power density level could be considerably greater than this if it were desired, particularly if the diodes were in good contact with a conductive substrate.

Rectenna cost. Rectenna cost per unit area will increase sharply, which of course favors using the device at its highest normal power density.

2.6.3 Significance of the Consideration of Work at Even Higher Frequencies Upon the Direction of the 20 GHz Experimental Program

The tendency of a high frequency scaled device to have a high packing density of diodes with a small junction area argues strongly toward an eventual monolithic approach in which the diodes and circuits are produced on the same substrate. This means a substrate of either gallium arsenide or silicon. This was an argument toward basing the 20 GHz work, upon a material whose dielectric constant is close to that of silicon or gallium arsenide. Alumina is such a material.

2.6.4 Discussion of the Different Approaches to a 20 GHz Rectenna

This section consists of the discussion of three approaches: (1) the direct scale of the thin-film, etched-circuit, 2.45 GHz rectenna; (2) use of the interdigital finger approach and the pedestal support which allows the complete

circuit to be laid down on one side of the dielectric substrate; and (3) the use of an alumina ceramic as the circuit substrate that fills the void between the foreplane and the reflecting plane of the rectenna and upon which the microwave circuit is printed. Conclusions and recommendations are then made in Section 2.6.6.

2.6.4.1 Direct Scale of the Thin-Film, Etched-Circuit 2.45 GHz Rectenna

Scaling the present structure is attractive in some respects but is unattractive in several others. One problem is that in a true scale the dielectric substrate would have to be decreased in thickness by the scale factor of about eight. Because of the lack of bonding materials (that bond the copper to the dielectric film) that are both very thin and of acceptable dielectric loss, it would be impractical to reduce the present sandwiched material from 2 mils to one quarter mil.

To compensate for the inability to scale the thickness, the capacitors in the low pass filter circuit section can be increased in area from the normal scale size but then they become quite large. It appears to be impactical to scale the shorting capacitance in the rectifier circuit to a larger area than the scaled size so that the capacitor would lose much of its effectiveness as a low impedance shunting circuit.

2.6.4.1.1 Application of Commercial Process for Vaporizing Metal on Thin-Film. One way to cope with the scaled thickness problem is to use a different process for putting metal on a dielectric film. One approach is to vaporize a thin film of metal onto a thin dielectric film in vacuum. This general approach is commercially used in depositing a thin aluminum film on one or both sides of a dielectric film for the diverse uses of thermal insulation and food preservation. Such material is available in the form of aluminum deposited on Kapton only 0.3 mil thick, and on mylar only 0.25 mil thick.

The thickness of the metal film is, of course, very important in any potential microwave rectenna application, both from the viewpoint of conducting heat away from the diode to a radiating surface, and from the viewpoint of electrical conductivity. From the conductivity point alone, the deposit should be at least equal to the skin thickness which, at 20 GHz and for aluminum, is 57 microns. However, the commercial material is only available with a maximum thickness of about 8 microns. The material is put on with a thickness of about 1.5 microns per pass, so that it would take about forty passes to establish the minimum thickness desired. But a more important limitation is that, after a few successive passes, the adhesion between the aluminum and the dielectric film becomes poor. The reasons for this is not clearly understood, but because the present commercial markets do not need a thick film, there apparently has not been a substantial effort to develop an understanding.

From a technical point of view, it is relatively straightforward to deposit copper on a thin film in the same way that aluminum is deposited. However, it is not done commercially because no need has developed to do so.

One of the interesting possibilities to build up the thickness of the metal on the side of the film on which the dipoles and bus bars are etched is to first etch the microwave pattern on the metalized film and then to use the bus bar circuits to conduct current into an electroplating bath to build up the thickness of the microwave pattern to the desired thickness. If the plating operation were technically feasible, and it would be certainly so if the film first deposited was copper, the process would be a very economical way to produce a rectenna if the microwave circuits were all on one side of the film. There is, of course, no way to contact the patches of metalized material on the other side of the dielectric film that form the microwave capacitors for the low-pass filters. For this reason, the technology of placing the circuitry on one side of the dielectric film that is discussed in Section 2.2.4.2 is of interest. However, the resistive loss in the metal patches as laid down by the vaporization process may be acceptable under certain circumstances, but more study would be needed, possibly in the process of investigating the losses in Kapton at 20 GHz.

Going to this thin film makes it possible to design an adequate shunting capacitor in the rectifier circuit and it may not be necessary to put a patch on the other side of the film for this capacitor if the supporting pedestal approach discussed in the following section is used.

2.6.4.1.2 Application of the Pedestal Support Technique for Better Heat Dissipation and for Separating the Foreplane from the Reflecting Plane. A technique for improving the heat dissipation capability of the rectenna was discussed in the proposal. The suggested technique was to bond the electrically neutral patch associated with the shorting capacitor to a metallic projection or pedestal from the reflecting plane. The outstanding advantage of this technique is that it "short circuits" the heat flow that would normally be along the transmission line and into the antenna dipoles for radiation. Instead, the heat is conducted efficiently into the ground plane where the entire ground plane can be used for effective heat radiation.

Even though the patch is separated from the conductor that is directly attached to the diode by two mils of Kapton F whose heat conductivity is only 4×10^{-4} calories per second, per centimeter, per degree C, the thermal resistance in the 2.45 GHz case is only 10°C per watt of transmitted heat--considerably lower than the drop between the diode and the top of the capacitor. A direct scale to 20 GHz would increase this resistance to 80°C per watt of transmitted heat.

It is easy to visualize a reflecting plane with many support pedestals about 0.100 inch high that have been precision milled as a monolithic structure. The pedestals will now be so close together as to function as the sole support for the rectenna foreplane.

From a microwave circuit point of view, it is possible to employ this technique because of the balanced nature of the microwave circuit. The patch

that represents an essential part of the shorting capacitor in the rectifier circuit is at neutral or zero microwave potential and is not connected to the DC circuits. Hence, placing it in full metallic contact with pedestals from the ground plane does not impact the DC circuits and theoretically does not impact the microwave behavior of the device.

To experimentally check this assumption, the capacitor patch was bonded to the end of a one-half inch diameter aluminum rod whose other end was attached to the cover plane of the standard test fixture (see Figures 2-1 to 2-3). The cover plate serves as the reflecting plane of the rectenna. The test results, over a wide range of incident microwave power and DC load resistance, were nearly identical to the test made on the same element in which the capacitor patch was separated from the reflecting plane by two centimeters of air.

2.6.4.1.3 Registration of Microwave Circuit Patterns on Opposite Sides of the Dielectric. Although there is now no trouble in getting registration between the pattersn on the opposite sides of the thin film at 2.45 GHz, it may be difficult to achieve registration at 20 GHz. But this does not mean that it cannot be accomplished.

2.6.4.1.4 Bonding the Beam Lead Diode to the Circuit. Normally, beam lead diodes are attached to circuits by thermal compression bonding. But the substrate is usually a hard material such as ceramic. The experience that the writer has had is that the Kapton is a soft material compared to ceramic and that the necessary pressure to make the bond pushes the copper circuit to which the beam lead is being welded into the Kapton. This is an area that would have to be studied further before proceeding with this general approach.

2.6.4.2 Use of the Interdigital Finger Approach to Forming Capacitors that Allows the Complete Circuit to be in One Plane

Figure 2-26 shows what the resulting circuit, except for the large shorting capacitor in the rectifier circuit, would look like if the interdigital capacitor technique were applied to the existing rectenna circuit at 2.45 GHz. One of the advantages of the etching of the microwave circuits in a single plane is that very thin dielectric film with vaporized metal coating on one surface only and as described in Section 2.6.4.1.1 may be used. Such a technology is also desirable for a monolithic approach to the construction of the rectenna element, as will be discussed in Section 2.6.4.3. And some of the objections to using the technique on a very thin film disappear when the interdigital fingers are placed on a thick substrate with a high dielectric constant, as would be the case in a monolithic structure.

2.6.4.2.1 Procedure for Designing Capacitors Formed by Interdigital Fingers. The design procedure for microwave capacitors formed by interdigital fingers is based upon the use of a formula for capacitance between many parallel flat strips of infinitely thin metal, with alternate strips constituting two sets,

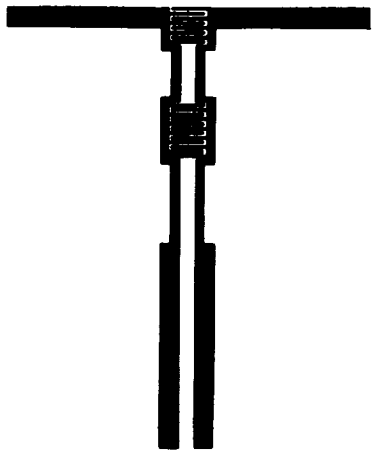

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Figure 2-26. Application of Intedigital Finger Technique to Form Capacitors on One Plane as Applied to a Rectenna Element at 2.45 GHz.

each set conductively tied together and representing one side of the capacitor. The capacitance per unit length between any two of these fingers is as follows.

$$C = \frac{\pi\varepsilon}{2 \ln \left[2 \cot \frac{\pi}{4} \frac{f}{f + d}\right]} \text{ Farads per centimeter}$$

where:

$$\varepsilon = \frac{1}{36\pi} \times 10^{-11} \text{ Farads per centimeter} \quad (1)$$

d = spacing between finger edges

f = width of each finger

This formula was experimentally checked out by measuring the capacitance between a set of 14 interdigital fingers, each 5 cm long, that were made from thin 0.40 cm wide copper strips spaced the same distance (0.40 cm) from each other on a thin piece of mylar. It was assumed that the mylar would have negligible impact upon the capacitance measurements. The measured value of the set of interdigital fingers was 5.5×10^{-12} Farad, which compares with a calculated value from the formula of 5.72×10^{-12} Farad.

The design of the capacitor then consists of a compromise between the number of fingers, their length, and their separation to provide the desired value of capacitance, using the expression (1).

2.6.4.2.2 Application of Interdigital Capacitor to Thin Film Rectennas. The difficulty in applying this technique to interdigital fingers on film so thin that it is assumed that the capacitance between fingers is not appreciably affected by the higher dielectric constant of the film is that the area taken up by the capacitor either has to be sizeable or the separation between fingers has to be very low. In Figure 2-26, the separation between fingers is 15 mils or 0.375 millimeter. This would scale to approximately 2 mils between fingers at 20 GHz.

The microwave shorting capacitor used in the rectifier circuit presents a difficult problem because the interdigital finger approach would require ten times the area of the capacitors for the low pass filter and would not therefore be practical. However, if the support pedestal discussed in Section 2.6.4.1.2 is used, then there may be no need for an interdigital capacitance in the rectifier circuit. The problem then, without a metal patch on the back, will be to position the pedestal so that the proper tuning of the rectifier circuit results.

2.6.4.2.3 Application of Interdigital Capacitor to Circuits on a Ceramic Substrate. If the interdigital fingers were completely immersed in dielectric material with a dielectric constant greater than unity, the capacity per unit length, as given by expression (1) in Section 2.6.4.2.1, would increase by the dielectric constant. If the material is alumina, the increase is a factor of 8.8. However, if the fingers are printed on top of an infinitely thick ceramic, the increase will be less than this. It cannot be assumed that the increase will be exactly half of the 8.8 or 4.4.

The capacitance between infinitely thin interdigital fingers on an infinitely thick slab of high dielectric constant material is undoubtedly an interesting theoretical problem. However, for design purposes, the increase can be experimentally measured. After mounting the set of interdigital fingers, used to check the expression for capacitance (in Section 2.6.4.2.1) on an alumina ceramic substrate, it was found that the capacitance had been increased by a factor of 5.22. Hence, the effective dielectric constant is 5.22 instead of 4.4.

The higher dielectric constant makes the use of the interdigital finger approach to a single plane circuit much more practical. The filter capacitances can be made without difficulty. And it is reasonable to expect that a solution can be found for the capacitor in the rectifier circuit with the much larger capacitance between a given set of fingers.

2.6.5 Use of Alumina Ceramic as a Microwave Circuit Substrate and as a Filler Between Foreplane and Reflecting Plane

This section discusses the advantages and disadvantages of the use of alumina ceramic, and in particular discusses the great advantage of transferring heat from the diode to a large heat sink. The increase in heat transfer is so dramatic that it is discussed separately in Section 2.6.5.2.

2.6.5.1 General Listing of Advantages and Disadvantages to the Use of Alumina as a Substrate

There are many advantages and comparatively few disadvantages in the use of alumina ceramic as a substrate. The advantages are:

- It eliminates an assembly operation of the foreplane to a supporting structure—certain to be difficult because of the small dimensions involved.
- It provides a hard surface for the thermal compression bonding of beam lead diodes to the structure.
- It provides an excellent low resistance thermal path for the dissipated power from the diode to flow to the reflecting plane which can be highly thermally conducting and treated to radiate heat efficiently. This is treated separately in Section 2.6.5.2.

- The high dielectric constant of an alumina substrate greatly increases the capacitance between fingers of an interdigital capacitor and makes their physical realization much easier.

- The microwave losses in pure alumina are quite low. Alumina in nearly pure form but with enough glass in it to make it vacuum tight has been satisfactorily used as windows in high powered microwave tubes for many years.

- The dielectric constant of alumina, 9, is close to that of silicon (11.8) and gallium arsenide (10.9). Such semiconductor material would presumably be used in a monolithic construction in which the diodes were constructed on the substrate. Using a ceramic substrate is a good intermediate step toward a monolithic structure.

- There is a large base of both thick and thin film circuit technology associated with ceramic substrates.

- The resulting structure is very compact--only one millimeter or less in thickness.

The disadvantages of an alumina substrate appear to be minimal. One problem is related to single plane circuitry, that of designing enough capacitance into the microwave shorting capacitor that tunes the rectifier circuit. However, the use of a high dielectric substrate makes it less of a problem. What is needed in this case is an investigation of the effectiveness of a very large number of interdigital fingers distributed over a substantial length of line. This will not look like a lumped capacitance at microwave frequencies but could possibly be as effective as a lumped capacitance. Even the present arrangement in the 2.45 GHz rectenna is not a lumped capacitance in the microwave sense.

Another possible disadvantage is the increased density of the elements. In a truly monolithic circuit with the diodes formed on the substrate, this would not be a disadvantage. And the power handling capability of the rectenna will increased with the packing density of the elements. It does become a disadvantage if it were desired to operate the rectenna at a low power density and to spread it over a large area, particularly if the diodes were separately constructed.

The ceramic construction might initially appear to be more expensive than other forms, and it certainly is at low frequencies, but it may well be the most economical at 20 GHz and above because it eliminates several difficult alignment and assembly steps that would have to be taken with other technologies.

2.6.5.2 Discussion of the Alumina Substrate as a Conductive Cooling Mechanism for the Diode

The bonding of the diode to a metal surface, probably gold, immediately above the alumina provides a cooling mechanism that is basically different from the approach currently used in cooling the 2.45 GHz thin-film, etched circuit rectenna, and even the pedestal technique (discussed in Section 2.6.4.1.2) where the heat still has to flow through a section of the transmission line before reaching the pedestal. When a ceramic substrate is used, the heat can flow directly from the diode to the reflecting plane where it can be radiated or transferred to a convective coolant.

Because alumina ceramic is a good heat conductor, 0.055 calories/degree C/cm/sec, and the space between the foreplane and the reflecting plane is only about 1 millimeter for a 20 GHz rectenna, the resistance to heat flow from each diode to the reflecting plane is relatively low. This fact, along with the high packing density of the diodes, allows several kilowatts of heat to be transferred for each square meter of surface area of the rectenna.

This conclusion may be supported by a simple mathematical model of heat flow from the heat-sinked diode to the reflecting plane. The model assumes that heat flows through a 90° truncated cone from the diode to the reflecting plane. The quantitative results of the use of this model are expected to be close to those for a much more complicated model that would simulate the real situation.

The resulting expression is:

$$W = (0.72) \frac{x_1 x_2}{x_2 - x_1} (\Delta T) \qquad (1)$$

Where

- W is the heat flow in watts from each diode.
- ΔT is the temperature difference between x_1 and x_2.
- x is the distance from the tip of the cone.
- x_1 is the position of the heat sink of the diode, and because of the 90°C cone assumed, is also the radius of the heat sink.
- x_2 is the distance of the reflecting plane from the tip of the cone, and is equal to the thickness of the ceramic plus x_1.

If the assumption is made that the radius x_1 of the heat sink of the diode is 0.00125 cm and the thickness of the ceramic is one millimeter, x_1 and x_2 become 0.00125 cm and 0.10125 cm respectively. When these are inserted into expression (1)

$$W = 0.0009 \Delta T \quad (2)$$

If ΔT = 100°C, certainly a conservative value, W becomes 0.09 watts for each diode. As indicated in Section 2.6.2, there could be 30,000 diodes per square meter in a 20 GHz rectenna. Thus, there could be 2,700 watts of heat flow for each square meter of rectenna. If the rectenna were only 60% efficient, it could handle 4,050 watts of power per square meter. Semiconductor substrates have a higher coefficient of heat transfer. For the same dimensional parameters and efficiency 12.5 kW/m^2 could be obtained with GaAs (see Section 2.7.2.2).

However, the reflecting plane to which the heat is transferred would have to be convectively cooled with air or liquid, since its radiation capability for a scenario of 150°C surface temperature, 30° ambient temperature, and emissivity of 0.5 is only 669 watts for one side and 1338 watts for both sides.

2.6.6 Conclusions and Recommendations

A firm conclusion resulting from this task was that the 20 GHz rectenna should use a solid dielectric separator between the rectenna foreplane and the reflecting plane. Two different constructions were considered. One was a hybrid construction consisting of an alumina substrate, silk screened circuits, and beam lead diodes. The other was a monolithic construction utilizing a semiconductor substrate. But because of the anticipation of actually constructing at least one rectenna section, with the only possible available form of diode being the beam lead diode, the hybrid construction was favored.

This approach proceeded to the point of making a search for beam lead diodes and procurement of ceramic substrates. However, as an appreciation of the difficulty of the task of putting just one element together grew, the enormity of the task of building a large rectenna area with the hybrid technology also grew. As indicated in the introduction to the next section (2.7) the monolithic construction turned out to be the recommended approach even though the recommended material GaAs was still an imperfect material. However, most of the work in Section 2.6 is applicable to the monolithic construction, including the interdigital approach to a single surface microwave circuit and the determination of diode dissipation capabilities.

2.7 Preliminary K-Band Rectenna Design (Task 7)

2.7.1 Introduction and Summary

The investigation of various approaches to the design and construction of a 20 GHz rectenna carried out in Section 2.6 (Task 6) found that the best approach for a non-monolithic structure would be a ceramic substrate with the microwave circuits silk screened upon it. The capacitors in the low pass filter circuit would be interdigital as shown in Figure 2.6.2 and as discussed in Section 2.6.4.2.3. The diodes would be beam lead diodes bonded to the silk screened gold transmission lines by thermal compression. The metallic reflecting plane would be deposited on the back side of the ceramic substrate. One remaining problem would be the bypass capacitor which would be of the order of one picofarad and could not be built in the interdigital format.

It was felt, however, because of the very large number of diodes that would be needed per unit area, that a completely monolithic structure would probably be the eventual answer with the diode fabricated on a silicon or GaAs substrate. In fact, at the time of writing the final report, GaAs monolithic technology was moving rapidly enough to perhaps justify the bypassing of the non-monolithic technology altogther if only a few rectenna elements were to be constructed for evaluation and demonstration purposes.

When the study in Section 2.6 was performed there was the hope of actually fabricating one or two non-monolithic rectenna elements including diodes at 20 GHz, and testing them during Task 2.7. That objective, together with the state of monolithic technology, less mature than now, encouraged the study in Task 6 focus on non-monolithic technology. That situation has changed.

There has also been the tendency to steer away from monolithic technology because of its perceived high cost. But it should be recognized that rectennas made at 20 GHz and higher with any technology will be expensive. Monolithic technology may not only be better than non-monolithic technology at 20 GHz but it is the only approach at frequencies significantly above 20 GHz.

The next section of this report, 2.7.2, will contain approaches to the fabrication of the monolithic rectenna as suggested by the Raytheon Research Division, who currently are fabricating what they feel are much more complex monolithic circuits than that represented by the rectenna. According to them, it would be quite feasible to fabricate a few rectenna sections on a selected area of a GaAs wafer. However, the quality of the material throughout a 3 inch GaAs wafer is currently not good enough to utilize the whole wafer without probably having an appreciable percentage of defective diodes.

Section 2.7.2 will be followed by Section 2.7.3 which is devoted to a scaling of the antenna portion of the K-band rectenna element as designed for an alumina or GaAs substrate to 2.45 GHz where extensive cold tests were made

to match the rectenna structure to space. From this work, the length of the rectenna dipole and the thickness of the ceramic substrate were determined.

The match was then checked out in "hot test" with the balance of the thin-film rectenna element in the 2.45 GHz exapnded waveguide test fixture. The "hybrid" element was found to perform at high rectification efficiency and with an acceptable amount of reflected microwave power.

In concluding the summary aspect of the introductory section it is necessary to point out that the rectenna design presented here is the first iteration of the design. During this first iteration unexpected inputs occurred when the individual rectenna element was integrated into the complete rectenna. This was caused in part by the early emphasis upon attempting to make a rectenna element for test, and in part by the successfully used procedure in the past to concentrate upon the individual element development before consolidating it into the rectenna.

The unexpected difficulty encountered in this instance was the bus bar loss caused by the much smaller cross section because of the scaling and the generally higher power density of the DC power output. To keep the bus bar power losses to within reasonable limits, 2 to 5%, the thickness of the bus bar would have to be 2 to 3 mils thick. This would undoubtedly cause difficulties first in building up such a layer on the gallium arsenide substrate and then in the mechanical sheer stresses at the interface caused by the difference in the coefficient of expansion. The coefficient of expansion of GaAs is $5.9 \times 10^{-6}/°C$ while copper and gold are $9\text{-}10 \times 10^{-6}/°C$.

Fortunately, there may be a solution to the problem by connecting the DC outputs of the elements in series rather than in parallel. There is some precedent for this in that the rectenna elements of early successful rectenna were connected in series. However, this possible solution would have to be investigated in more detail, particularly in the context of rectenna stability.

Another element that was recognized later in the study was that the sections of transmission line that are needed as the inductive element in the low pass filter have a low characteristic impedance because they are laid down on a material with a high dielectric constant. This low characteristic impedance makes it difficult to make a low pass filter with a sufficiently high characteristic impedance to match the rectenna element into a high value of DC load resistance, for example, 200-400 ohms. The indication then is a lower resistive load, perhaps in the 100 ohm range.

It is difficult to predict the results of a reiteration of design that takes a lower impedance level into account. There are benefits as well as non-benefits. A low characteristic impedance means that the C_{to} of the diode can be made larger, thereby increasing the capacitance of the diode and reducing the series resistance and increasing the thermal dissipation of the diode-both

benefits. On the other hand, the DC voltage output will be less so that the Schottky barrier losses become more important. It can be concluded, however, that operation at a high power density level is highly desirable for efficiency purposes.

2.7.2 Conceptual Design of a Monolithic Rectenna at 20 GHz

The design of the physical layout of a monolithic rectenna at 20 GHz and higher frequencies consists of two part. The first part is making the diode on the semiconductor substrate so that it can be connected properly to the microwave circuit. The second part is the layout and construction of the microwave circuit. We will review the design of the diode first.

2.7.2.1 Diode Design for 20 GHz Monolithic Construction

The diode design will be reviewed from the viewpoints of fabrication procedure, specification of design parameters such as C_{to}, reverse breakdown voltage, etc., and dissipation capability. An extrapolation will also be made to higher frequencies.

Fabrication Procedure

The construction of a Schottky barrier diode by epitaxial growth and metallization on a semiconductor substrate base is shown in Figure 2-27. It is assumed that the substrate is Gallium Arsenide but the construction would be similar for silicon. The first thing that is done is to lay down as highly a doped layer, "N"+, as possible by epitaxial deposit on the substrate to serve as a low resistance conduction path. Then an epitaxial layer of "N" GaAs that is doped in the proper amount and to the allowable minimum thickness is set down. Then small areas of "resist" (to resist etching) that correspond to the appropriate area of the Schottky barrier interface are laid down and the underlying structure is etched away to include some of the N+ area. An upward projection of the diode like a "mesa" results and gives the "mesa diode" its name.

A metal contact is then deposited on top of the mesa to form the Schottky barrier itself. This metal is often platinum but it can be other material. Tungsten is a material that results in a lower voltage drop across the Schottky barrier and therefore improves the efficiency of the diode, particularly at low DC voltage output of the diode.

The diode itself is now complete but it will be necessary to make a contact between the metal side of the Schottky barrier and one side of the microwave circuit, and to make a contact between the back side of the diode and the other side of the microwave circuit.

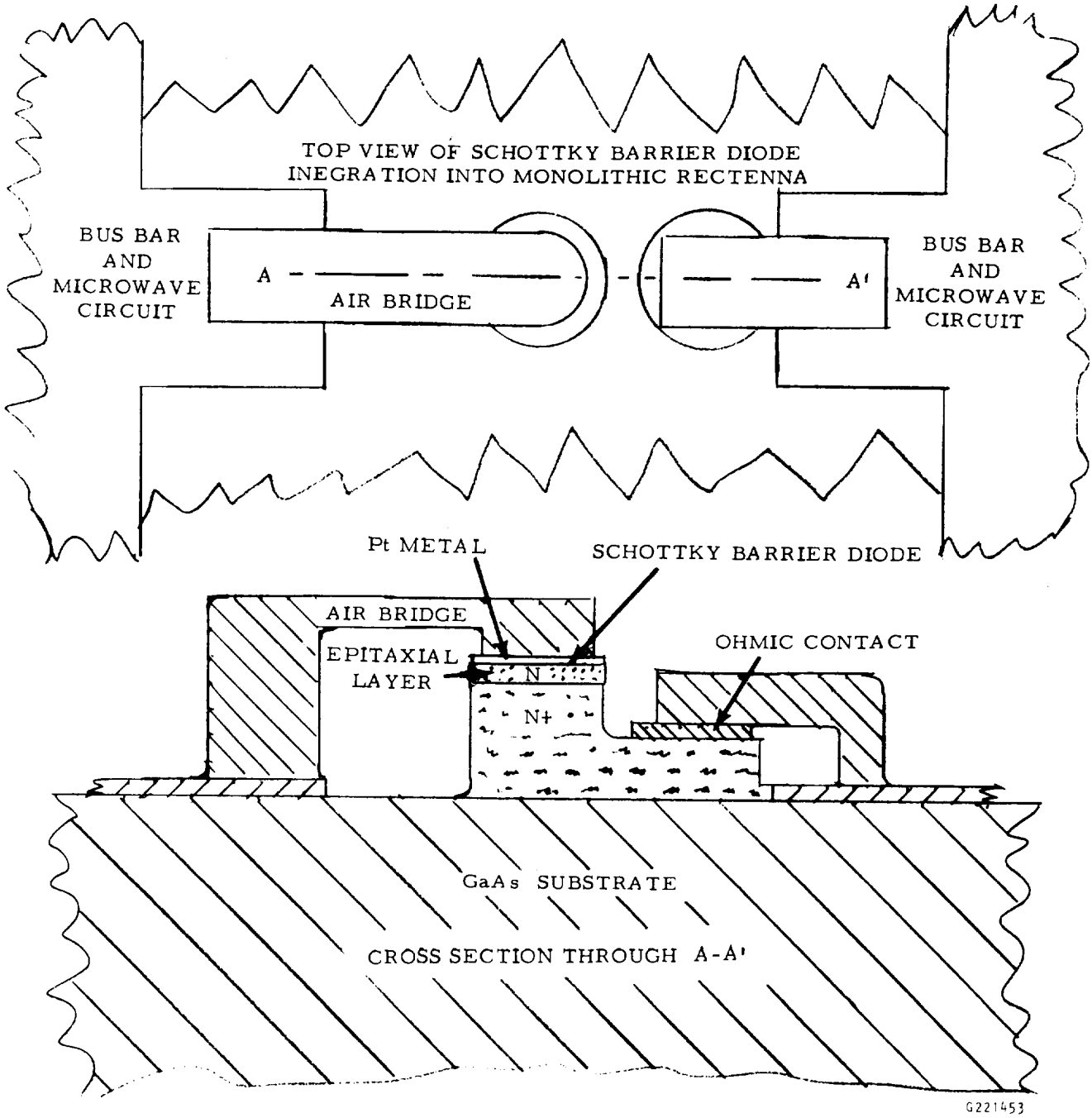

Figure 2-27. Diagram Illustrating the Schottky Barrier Diode Rectifier Portion of the GaAs Monolithic Structure for a Rectenna Element.

The contact with the top of the mesa diode is accomplished by a technique known as air bridging. Although other methods of contact could be used, and would be used in other approaches to the design of the diode, the air bridge has a unique advantage. Early in the development of the rectenna the ability of a diode short to disable the rectenna unless there was the equivalent of a fuse in series with it was recognized and fuses have been built into the packaged diodes since. In the monolithic rectenna, many diodes are operated in parallel across the DC bus. If a short occurs within one of the diodes, the short-circuit currents from the other diodes pass through the air bridge which will be very small in cross section and will act as a fuse. After the air bridge is burned through the rest of the rectenna elements return to normal operation.

From Figure 2-27 it is noted that the other side of the diode is connected to the other side of the microwave circuit through the N+ GaAs and an ohmic contact. The ohmic contact can be processed by applying "resist" to all the surface except the point at which the ohmic contact is to be made.

To reduce the resistance to current flow through the N+ material it is doped as much as possible to a level of 10^{18}, giving it a resistivity of 0.002 ohm cm. Although the path to the back contact is fairly long, the current has ample opportunity to spread out and this action coupled with the low resistivity makes the diode design approach acceptable.

Selecting the Initial Diode Design Parameters to Determine Internal Losses and Power Handling Capability

It is possible to initially select, largely on the basis of experience, design parameters of the diode from which it is possible to predict typical internal losses and the power handling capabilities of the diode. A final diode design would depend upon a number of reiterations that would probably be associated with the application of the rectenna and the environment in which it would be operated.

At the level with which we will deal with the design it is adequate to characterize the diode as an ideal rectifier with a capacitance in shunt with it, and a resistance in series with the combined diode and capacitance. The value of the capacitance varies with the potential across it, and it is typical to define this capacitance as that value it takes on when there is no voltage across it. This is called C_{to}, or the zero-bias capacitance.

The series resistance R_s is very important in determining the efficiency of the diode. In the conduction portion of the rectification cycle the DC current flows through it, while on the non-conducting portion the charging current to the capacitance C_{to} flows through it. The value of R_s depends upon the semiconductor material used, the doping density of the epitaxial layer, the thickness of the epitaxial layer, and the area of the Schottky barrier junction.

Another important parameter is that value of inverse voltage applied to the diode at which the diode starts to conduct in the reverse direction. This value of voltage is denoted V_b and its value is typically 2.5 times the DC output voltage of the rectifier, to allow for a peak inverse voltage also 2.5 times the DC output voltage. The parameter V_b determines the doping density of the epitaxial layer and its minimum thickness, and thus, in combination with the area of the junction, the series resistance R_s of the diode.

The first step in the design procedure was to select a value of C_{to} that is based upon retaining the microwave circuit impedance level of the 2.45 GHz design, and to scale the capacitance C_{to} by a factor equal to the ratio of the frequencies which is a factor of approximately 8. The C_{to} of the 2.45 MHz diode was 3 pf which implies a scaled value of 0.375 at 20 GHz.

As previously pointed out, another important parameter to specify initially is the reverse voltage breakdown, V_b, of the diode because this controls the doping density, the capacitance C_{to} per unit area of the junction, and the thickness of the epitaxial layer which along with the junction area controls the series resistance. It is assumed for an initial design that this voltage, V_b is 20 volts.

Given the reverse voltage breakdown of 20 volts, it is determined that the doping density should be 7×10^{16} atoms per cubic centimeter. (Figure 23, Chapter 5 of Reference 9.) For this doping density the zero bias capacitance will be about 70,000 pf/cm^2. (Figure 9, Chapter 3 of Reference 9.) The area of the Schottky barrier diode will then be 0.375/70,000 or 5.3×10^{-6} cm^2. This corresponds to the area of a circle with a diameter of 2.58×10^{-3} cm, or roughly 0.001 inch.

The thickness of the epitaxial layer may be obtained from the depletion layer which for a doping density of 7×10^{16} is 0.6 microns (Figure 25, Chapter 5 of Reference 9). Conservative design, however, would increase this to 1.0 micron thickness.

The resistivity of GaAs doped to 7×10^{16} is 0.08 ohm-cm. (Figure 22, Chapter 5, of Reference 9.) So for an epitaxial volume that is one micron thick (0.001 cm) and 0.00258 cm in diameter, the resistance R_s will be approximately 6 ohms. This resistance enters into the overall efficiency of the diode in complex ways but if the DC resistance level is in the range of 200 to 400 ohms, its impact on the overall efficiency will be a multiplying factor of about 0.9.

The comparable series resistance for silicon would be 30 ohms which would reduce the overall efficiency significantly. The situation will become even worse at higher frequencies because of the progressively smaller size of the junction area and no relief in the thickness of the epitaxial layer if the same breakdown voltage is to be maintained. It would therefore appear that the use of GaAs for high frequency rectennas is almost mandatory to retain a useable efficiency.

2.7.2.2 Dissipation Capabilities of the Diode and Its Power Handling Capability

The heat dissipation capability will now be examined with the aid of equation (2) in Section 2.6.5.2. Gallium arsenide with a heat conductivity of 0.64 watt/cm°C is a better heat conductor than is alumina so the first term in equation (2) become 2.0 for Gallium Arsenide. Now the expression (2) with X_2 much greater than X_1, as it is in the situation under discussion, is approximated by the term $2X_1$. X_1 for the 90° cone angle assumed for the heat flow is the same as the radius of the Schottky barrier junction which was 1.29×10^{-3} cm. Therefore the heat flow away from the junction, if the cooling is done only by heat flow to the backplane, is 0.0026 ΔT, where ΔT is the temperature difference between the diode and the back plate.

The determination of an expression for the approximate dissipation capability of the diode may now be used to approximately determine the DC output of the rectenna element in terms of the diode efficiency and ΔT. Although there are other losses in the rectenna element that should be taken into account in a more exact analysis, the losses in the diode are the predominant losses. The approximate expression becomes.

$$P_{dc} = \left(\frac{n}{1 - n}\right) \text{ (Power Dissipated in Diode)}$$

$$= \left(\frac{n}{1 - n}\right) (0.0026\ \Delta T$$

Where P_{dc} = DC power output from rectenna element

n = efficiency of diode

ΔT = temperature differential between diode and back plate

Figure 2-28 gives the contours of constant P_{dc} as a function of diode efficiency and ΔT.

From this figure it would appear that an expected DC power output in the range of 0.4 watts per diode element would be reasonable. What does this mean in terms of watts per square meter? On a strictly scaled basis the 200 elements per square meter in the 2.45 GHz array would scale the factor 8^2 to 12,800 elements. It is probable, however, that the elements will be much more densely packaged because of the high dielectric constant of the substrate. Thirty thousand elements would be more likely. Therefore, with a single rectenna power output of 0.4 watts, the power output per square meter would be 12.5 kilowatts.

This figure of 12.5 kilowatts assumes that the reflecting plane is conductively or covectively cooled in some manner. For an atmospheric application this is relatively easy to accomplish. However, for space applications

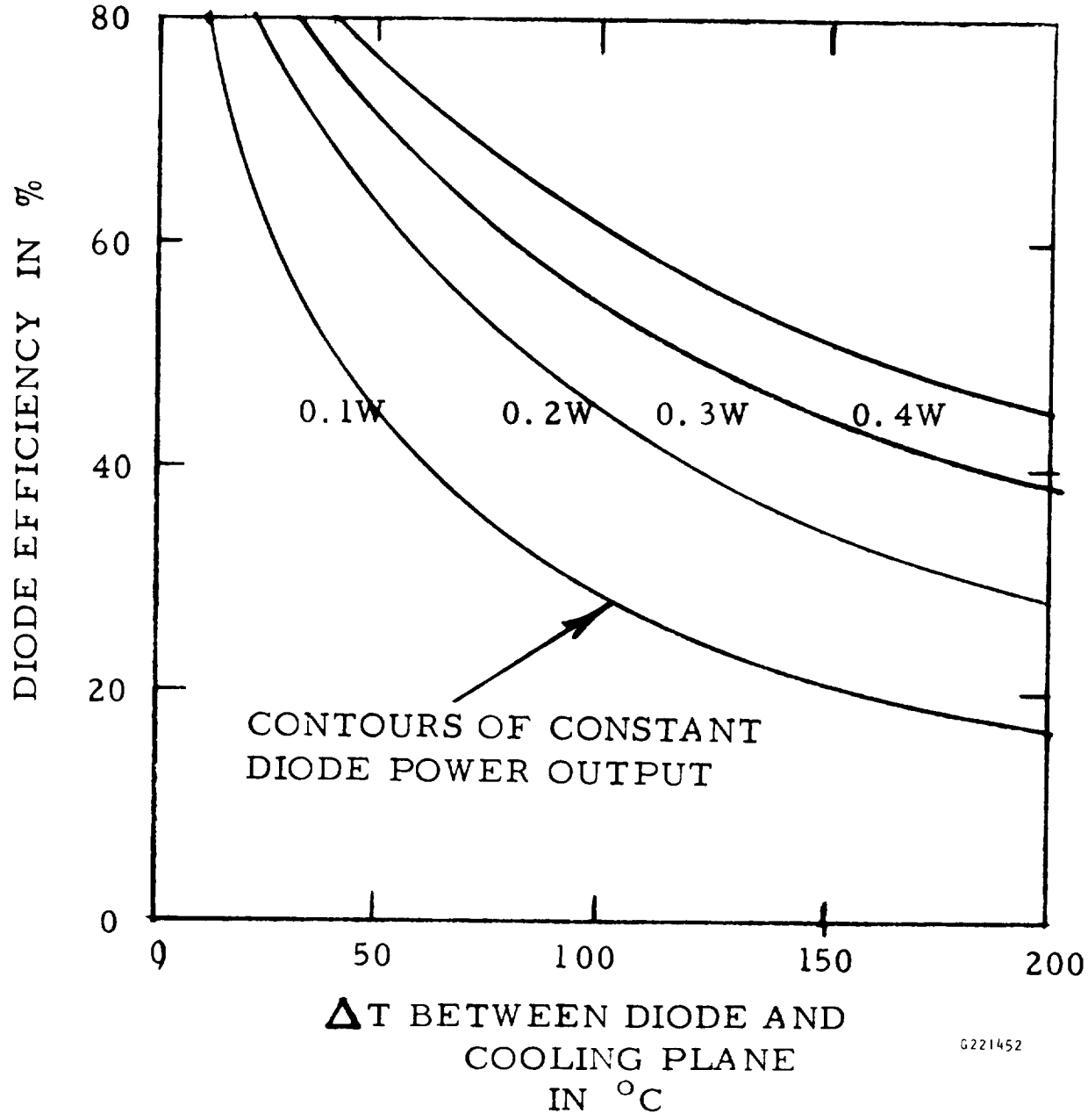

Figure 2-28. DC Power Output of Diode as Function of the Diode Efficiency and the Temperature Difference Between the Schottky Barrier of the Diode and the Reflecting Plane of the Rectenna Array. Assumptions are a GaAs Substrate, a Frequency of 20 GHz, and a Diode Spot Size of 0.00254 cm Diameter.

it may be necessary to rely upon heat radiation directly to space. If the reflecting plane is operated at 100°C, the maximum (black body) radiation to space from the back plane at a temperature of zero degrees Kelvin would be 1.09 kW. However, the front surface would probably be at nearly the same temperature, doubling the radiation to space. After making allowances for radiation into a higher ambient than 0°K and an emissivity of less than 1, radiated power in the range of 1 Kw/m^2 is more likely. This would seriously restrict the amount of DC power output from the rectenna. If the efficiency were 70%, the DC power output would be 2.33 Kw/m^2.

2.7.2.3 Output Voltage of the Rectenna Element

With a DC power output of 0.4 watts and a DC load resistance of 200 ohms, the DC output voltage would be 8.9 volts. The corresponding peak inverse voltage of 22.4 volts would exceed the reverse voltage breakdown of 20 volts established for the tentative initial design. However, with a limitation of 0.1 watt per diode which might well be the case for a radiation cooled rectenna, the DC voltage for a 200 ohm load would be only 4.5 volts.

2.7.2.4 Scaling to Higher Frequencies

Reasoning from the design sequence for the 20 GHz diode, it is reasonable to expect that the junction area will scale down directly with frequency, the internal resistance will scale up with frequency, and that the design dissipation capability of the diode will scale inversely as the square root of the frequency. However, the number of rectenna elements will scale up as the square of the frequency so that the number of diodes scales as the square of the frequency. The total dissipation of the diodes will then increase as the square root of the frequency. However, the series resistance is increasing to a significant value that will seriously impact the efficiency. Without a more detailed study and some experimental data for comparison purposes, the quantitative performance cannot be accurately predicted.

2.7.2.5 Microwave Circuit Design

The other part of the physical layout is the microwave circuit. It will be patterned after the technology discussed in Section 2.6.4.2 and Figure 2-26. It will be interdigitaled. Some study will have to be made on whether the interdigital microwave circuits are laid down on the substrate before or after the construction of the diodes, but this should not present a serious problem.

The minimum thickness of the films laid down as part of the microwave circuit is controlled by the skin thickness which is a function of the microwave frequency. At 20 GHz the skin depth in copper is 57 microns. A thickness double this or 114 microns would be adequate for microwave conduction even after allowing for current flow on both sides of the conductor. The corresponding thin film resistance (resistance to flow of microwave power) is

0.023 ohms per square. The conductors in the microwave circuit (per rectenna element) as scaled from 2.45 GHz are 0.3 cm long and 0.020 cm wide. There are about 15 squares in one of these conductors or 30 squares in the loop of two conductors. Because the current flows on both sides, the total microwave resistance is 0.69 ohms. If the conductors are of gold the resistance will be a little higher. Also the current will tend to accumulate on the inner surfaces of the conductors which will increase the resistance somewhat. But the values will be sufficiently low so that the rectenna element efficiency will not be seriously impacted.

The microwave conductors are also used as DC bus bars, to collect the rectified power. How thick will they need to be? The cross section of the bus bars and their length will determine the DC bus bar resistance and therefore bus bar losses. The resistance of the bus bars will be in series with the equivalent load resistor at the end of the row of diodes that are parallel together. If each rectenna element is to look into 200 ohms then the DC resistance across the bus bar will be 200 divided by the number of elements in parallel. The number in parallel on a 3 inch wafer could be typically 20, leading to a typical DC load resistance of 5 ohms. If it is desired to keep the bus bar losses to only 2% of the power output, then the equivalent resistance of the bus bar can be only 0.10 ohms. The loop length of the bus bar to service the 20 diodes, on the 3 inch wafer is 15 cm. If the bus bars average 0.04 cm in width, then it will be found that the thickness of a copper bus bar must be 0.003 cm, or 3000 microns, or 1.0 mil thick to keep the bus bar losses to 2%. This calculation reflects the reduced thickness needed because the full current is flowing in the bus bars only at the output end while no current is flowing at the input end, reflecting the fact that the diodes appear as generators evenly distributed along the length of the bus bars.

The 3000 micron thickness required for bus bar efficiency is 26 times the thickness needed to minimize microwave losses, and represents a challenge in the design and processing of the monolithic circuits. Of course, the bus bar thickness can be cut down by using fewer diodes in parallel, and by operating into a higher DC load resistance that in turn reflects the choice of a higher impedance level for the microwave circuit. But still the bus bar thickness will be much larger than that needed for the microwave circuit.

As explained in the summary in the introduction (Section 2.7.1) the necessity for heavy bus bars can be eliminated by connecting the elements in series, and indeed this has already been done in some of the early work on rectennas. From a monolithic device technology point of view this appears to be possible, and could have some additional advantages in eliminating the air bridge fuses if it is presumed that the failure mode of the diode is an internal short that would remain a short. Typical voltage outputs from 20 elements connected in series could range from 100 to 200 volts which would presumably be a generally convenient value of voltage to work with.

Also, as noted in the Introduction it may be difficult to maintain a high characteristic impedance of the low-pass filter sections at the input to the rectenna element. This situation will lead to modifications that will improve the dissipation capability of the diodes but at the expense of increasing the percentage of total power that is dissipated in the Schottky barrier. A combination of these considerations will force the rectenna power level upward to maintain efficiency.

2.7.3 Matching a Dipole that is Mounted on a Ceramic or Semiconductor Substrate

The previous sections have discussed both a monolithic rectenna with integration of the diode and microwave circuits and a hybrid rectenna that consists of an alumina substrate with silk screened circuits and separate beam lead diodes bonded to the circuits. In both of these approaches the necessity arises to match these structures to space so that the incoming microwave beam is effectively absorbed into the structure. This section discusses experimental efforts that were successful in matching an antenna on an alumina substrate to space. Because the dielectric constants of the semiconductor materials GaAs and Silicon are very close to that of alumina the results will be applicable to monolithic type rectennas.

Because it would be difficult to perform the experimental work at 20 GHz both because of test equipment considerations and because of the high precision required in laying out the physical parts of the experiment, the work was carried out at 2.45 GHz. However, the results can be readily scaled to any other frequency including those much higher than 20 GHz.

For these experimental measurements it was most desirable to use a Hewlett Packard network analyzer. The output of these analyzers, of course, is in the form of a coaxial output that is unbalanced with respect to ground while the rectenna dipole inputs are balanced with respect to ground. However, the rectenna circuit can be split in half by establishing a ground plane as illustrated in Figure 2-29. The center conductor of the HP network analyzer output is connected to one side of the dipole antenna while the outer conductor is connected to the ground plane. It is, of course, necessary to incorporate the reflecting plane of the rectenna into the set-up shown in Figure 2-29. The reflecting plane of the rectenna should not be confused with the ground plane of the measurement set up.

The varying input impedance of the radiating dipole as seen from the network analyzer terminals was obtained as a function of thickness of the ceramic substrate and the length of the dipole as shown in Figure 2-30. Conditions were found for a reflectionless match from a 100 ohm source. The 100 ohms corresponds to the 50 ohm impedance of the HP network analyzer source used in conjunction with the rectenna dipole element sliced in half with the use of the ground plane because the impedance level of the rectenna element is halved by this procedure. The match was obtained with an antenna length of

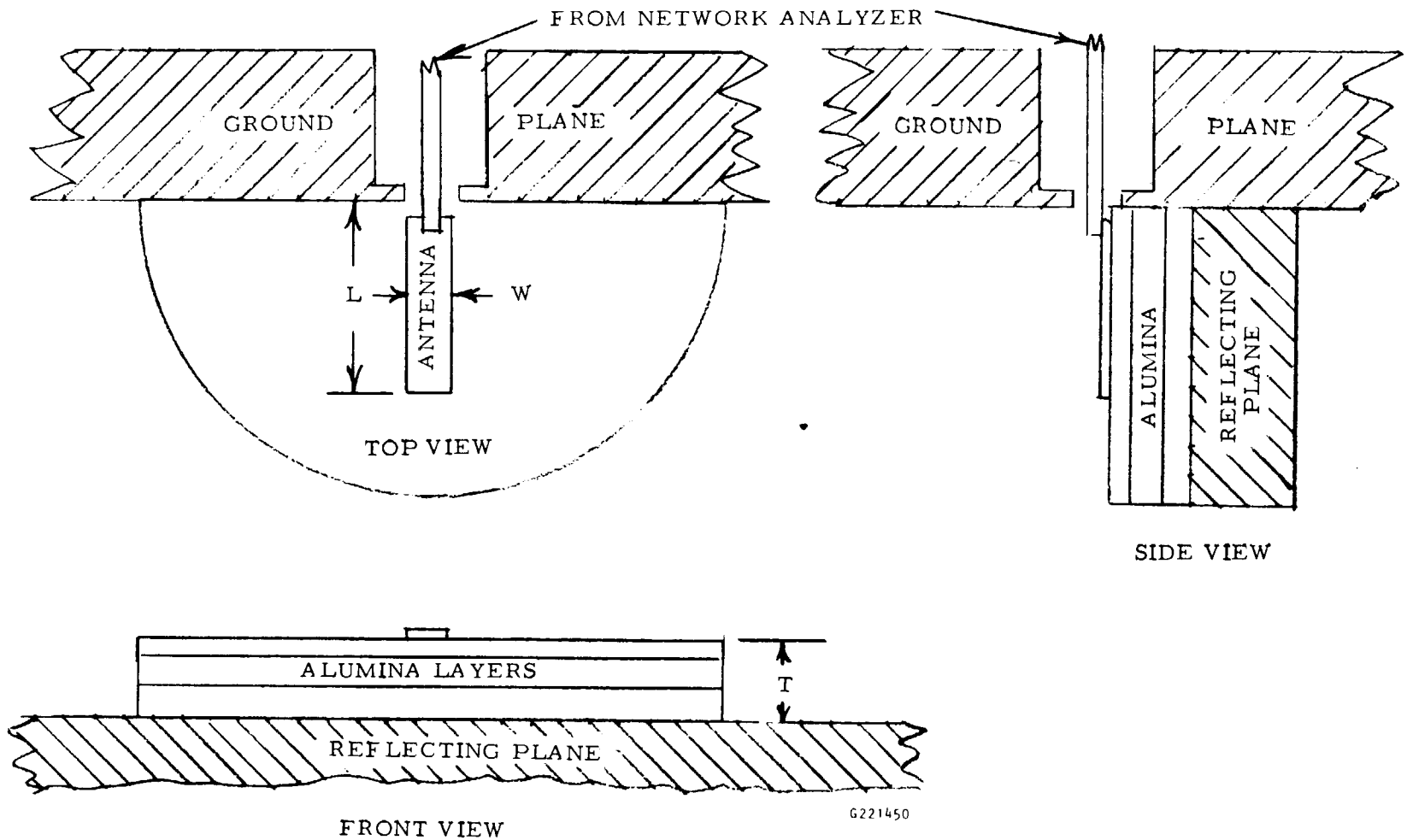

Figure 2-29. Test Arrangement for Matching a Rectenna Element Dipole that is Mounted on an Alumina Substrate to Space. The Rectenna Element is Split in Half by Means of a Ground Plane which Leaves the Characteristics of the Rectenna Element Unaffected Except for Halving the Characteristic Impedance. In this Manner, a Standard Network Analyzer can be Used to Make Measurements of the Match between the Element and Space.

OF POOR QUALITY

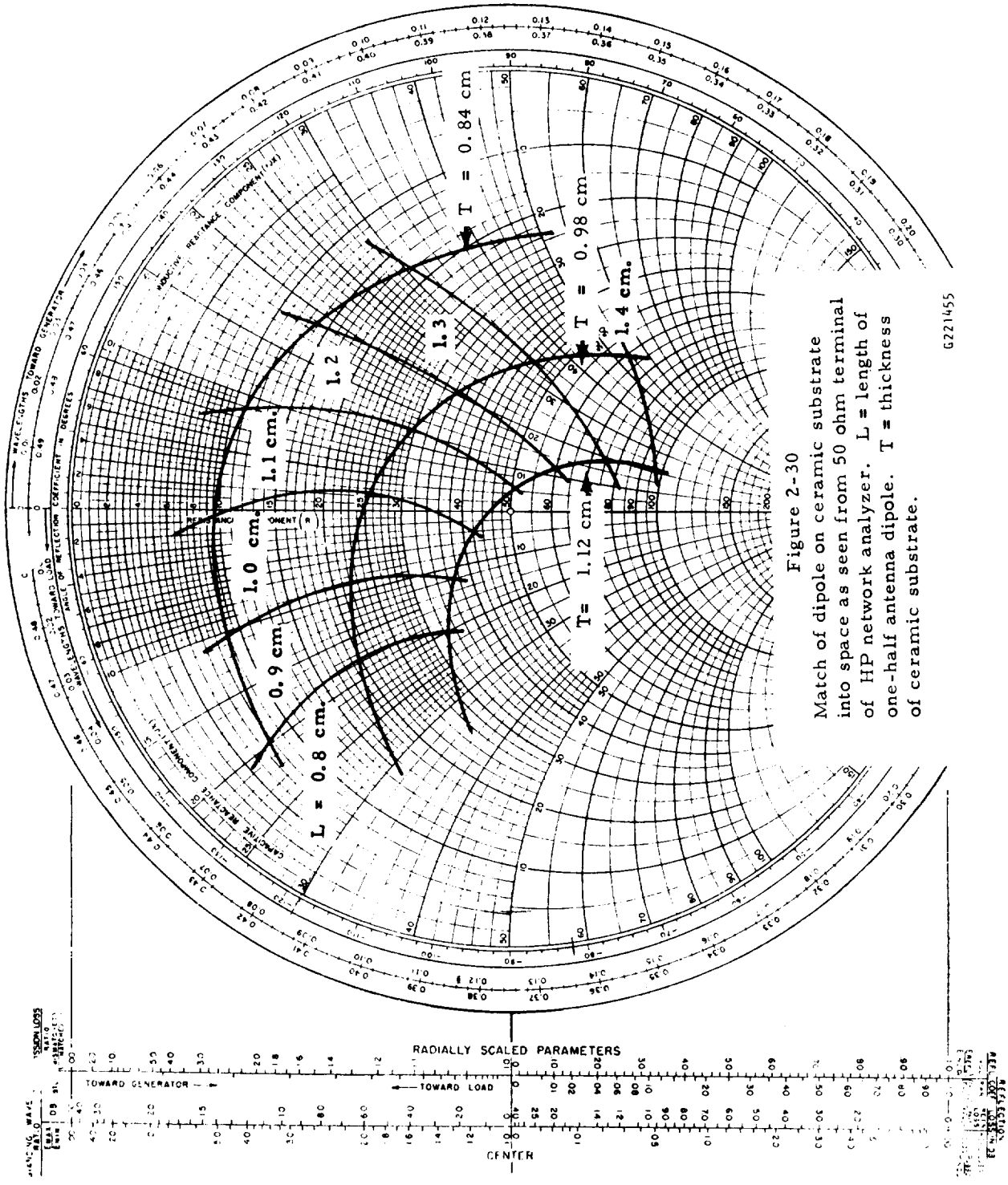

Figure 2-30

Match of dipole on ceramic substrate into space as seen from 50 ohm terminal of HP network analyzer. L = length of one-half antenna dipole. T = thickness of ceramic substrate.

1.05 cm and an alumina thickness of 1.12 cm. The length of the antenna at 2.45 GHz in the conventional thin-film, etched circuit rectenna described in Section 2.1 is 2.92 centimeters. The length is roughly shortened by a factor of 3, approximately the 1/2 power of the dielectric constant.

2.7.4 Measurements of Match, Power Output, and Operating Efficiency of a 2.45 GHz Rectenna Element Mounted on a Ceramic Substrate

Having obtained a match between space and an antenna dipole mounted on alumina substrate it was of interest to place a rectenna element with its dipole mounted on ceramic within a closed system so that measurements of match and overall efficiency could be made. A thin-film, etched-circuit rectenna element was used for this purpose. To prevent the ceramic material from modifying the low pass filters in the thin-film, etched-circuit format, this portion of the microwave circuit was elevated 0.100" above the ceramic. The antenna, shortened in length as required, was mounted directly on the ceramic.

It would be expected that the ceramic mounted dipole would have a cell area within the rectenna array considerably less than would a dipole separated from a reflecting plane with only air or vacuum in between. It would therefore need a test fixture with a smaller cross section to simulate the cell area. This was, indeed, found to be the case. Three differently configured sources of incident microwave power were used. The first was the standard expanded 4.5 x 4.5 inch fixture; the second was standard 4.5 x 2.25 inch waveguide, and the third was standard 3 x 1.5 inch waveguide. Waveguide openings had waveguide flanges so that there was a minimum of microwave leakage to space.

A best match of 4.8 dB (7.2 percent reflected power) was obtained with the 4 1/2 x 2 1/4 inch waveguide source. The resistive load was 300 ohms; the DC power output was 0.848 watt, and the element efficiency was 80.3 percent. For this performance the total length of the antenna dipole was 2.8 cm.

Before more work of this nature is done, it is recommended that the low pass networks and other portions of the microwave circuit also be placed on the ceramic substrate.

3.0 DISCUSSION OF RESULTS

A discussion of results is naturally divided into two parts. One part is that having to do with the results of the work on the thin-film, etched-circuit rectenna at 2.45 GHz. The other part is that having to do with the extension of rectenna technology to the frequency of 20 GHz and above.

3.1 Discussion of Results of the Program to Develop a 2.45 GHz Thin-Film Etched-Circuit

The broad intent of this program was to upgrade the rudimentary work on a thin-film, etched-circuit rectenna under a previous contract having to do with a rectenna for a microwave powered airship and to direct its further development toward use in space. The results of the program were highly successful in those aspects covered by the program. However, there were several aspects that were not covered.

The major achievement was the rather dramatic upgrading of the power handling capability of the thin-film rectenna for both space applications and for suitable terrestrial applications such as microwave powered aircraft. This upgrading in power handling capability resulted from several improvements. The first was an improvement in efficiency, from 70 to 85%. Because this increase in efficiency halved the inefficiency losses, it doubled the DC power output capability for a given rectenna operating temperature. Moreover, the change from mylar to Kapton substrate that caused the efficiency improvement also made it possible to operate the dielectric substrate safely at a temperature of 200°C where the GaAs diode can safely operate. Although difficult to quantify, it is expected that ability to operate at a higher temperature added another factor of 2 to the power handling capability of the rectenna in space or elsewhere.

It was possible to partially evaluate the power handling capability under incident microwave radiation up to one kilowatt of DC power output per square meter under conditions of convective air cooling at sea level air density. At this point, the high impedance design of the rectenna element limited further data taking. However, the diode temperature rise was still less than 100°C and the air flow velocity was nominally low. The implication is that the rectenna element should be redesigned to a much lower impedance level to accommodate higher levels of power density which might be desired for terrestrial applications and for which the rectennas is basically capable. In fact, going back to the impedance level of the rectenna design that was used so successfully in the 1975 tests on the Mojave desert appears to be desirable.

The scope of the contract did not include any testing in vacuum. However, from theoretical work based on the same radiative geometry (reported in reference 1) and then upgraded by the use of a ceramic package for the diode, it would appear that a DC power density of 400 w/m^2, twice that of the conservative

value previously assigned to a space application, would be reasonable. It is obvious that it is important to obtain data on the rectenna while it is being operated in a vacuum environment.

The rectenna that has been developed under this contract appears to basically meet requirements for aircraft use as far as efficiency, power-handling requirements, and life are concerned. There is, inevitably, some harmonic radiation from the rectenna which may limit its initial acceptance. And, recently, it has been observed that the current thin-film rectenna design has internal parametric oscillations at a few hundreds of MHz that modulate the 2.45 GHz signal reflected from the rectenna. Such sidebands in the reflected signal are intolerable. Current investigations indicate that this is not a general property of rectennas and that the phenomenon can probably be removed from the present design.

For space applications of the rectenna there are additional considerations having to do with the space environment that were not examined. The rectenna is impacted in space by various forms of radiation. The change to Kapton for the dielectric substrate eliminates deterioration from ultra violet radiation as a consideration, but the diode may be susceptible to other forms of radiation. If there is deterioration of the diode from such radiation, it could probably be minimized by adding shielding to the diode on a scale which would not add materially to the overall mass of the rectenna. The rectenna should be evaluated in an environment that closely simulates space in all respects, or be tested in the space environment itself.

There may also be the problem of leakage current flow through plasma between high voltage terminals of the rectenna when the rectenna is operated at relatively low altitudes. There is divided opinion as to whether or not that would be a problem with the current rectenna. However, if it is a problem in the current rectenna design, the leakage current could be minimized by adding an additional dielectric film over the etched copper circuits, much as is done with commercial flexible copper circuits which are bonded between two Kapton films. It is difficult to image a high voltage build up across terminals of the rectenna and causing arcs. Regardless of the polarity of the charge build-up, the diode serves as a means of allowing that charge to leak off in either the forward conduction cycle or by exceeding the breakdown voltage in the reverse direction.

There are a number of issues related to the fabrication and cost of the thin-film, etched-circuit rectenna. They relate to both the fabrication of the etched-circuit and to the diode.

For any large scale application the fabrication of the rectenna must be a continuous flow operation with the laminate coming in at one end and the finished etched rectenna coming out the other. The laminate must also be made continuously with the copper foil and dielectric substrate flowing into the

laminating machine from spools of the material. Initial continuous systems may be limited, for initial cost consideration, to a single rectenna row.

The Kapton F laminate itself requires a high temperature press which is non-standard equipment in most fabrication shops. What is needed is an adhesive that has a low dielectric loss but which will bond copper to Kapton H. In this context, the space application demands a dielectric substrate with minimal loss, but aircraft applications could tolerate nominal losses associated with Kapton itself and also nominal losses in the adhesive.

The diode that is currently being used is an exceptionally high quality diode in many aspects. The internal losses are low; it will operate at a relatively high temperature; and heat is conducted readily away through the plated heat sink. The diode has been evaluated on life test. If purchased to the proper set of specifications the diode will give satisfactory service for many years. The major issue with respect to the diode is its cost. Although it is a basically simple device, the very low demand for the diode has kept it at a high price. Downstream, at some point, there should be a new diode packaging technique which will make the diode easier to fabricate and also make it quite thin so that the rectenna can be rolled up tightly.

3.2 Discussion of the Results of the 20 GHz Rectenna Investigation

This portion of the study focussed upon investigating approaches to the design and fabrication of a rectenna at 20 GHz. However, the findings of the investigation were expected to be applicable to rectennas at considerable higher frequencies.

It appears that the investigation of a high frequency rectenna and preliminary findings on the best way to construct it may be timely. The recent development of the gyrotron, a new microwave tube that can continuously deliver several hundreds of kilowatts of power at these high frequencies, permits the generation of very sharp beams with a moderate sized aperture which could be mounted on ships or land vehicles. Thus, in principle at least, the high frequency rectenna and gyrotron expand the range of applications for free space power tranmission.

The investigation started out as a scale of the 2.45 GHz thin-film etched-circuit rectenna to 20 GHz. Various formats for the use of a thin dielectric film were examined in considerable detail. None, however, were without severe faults, so interest turned in the direction of using an alumina substrate and silk screened circuits - a technology in common use. Then the unavailability of suitable beam lead diodes, combined with the need for so many per unit area, and some concern about the difficulty of bonding the diodes to the circuits, motivated looking at the monolithic approach in which a semiconductor substrate is used and the diodes are built up on that substrate.

The monolithic approach was finally selected as the recommended approach for future development activity on high frequency rectennas. Further, GaAs was recommended over silicon as the substrate material because of its much better microwave properties, even though the quality of the substrate material needs upgrading, an activity that has now received priority because of the desire to use GaAs in other applications.

It was determined that a high frequency monolithic rectenna would require operation at a comparatively high power density to maintain an acceptable level of efficiency because of the Schottky barrier voltage becoming comparable to the output across the load and therefore representing a large loss. The power density for good efficiency will be in the 3 to 10 kilowatt per square meter range and the accompanying heat dissipation will be comparable, necessitating some form of convective cooling of the rectenna. This introduces a severe problem for space applications but one that can be tolerated for air borne applications where the higher power density may be wanted anyway.

Although the monolithic rectenna approach has been established as the direction in which to proceed with a high frequency rectenna development, it must be pointed out that the development will probably be a lengthy and costly one before production prototypes could become available. If such developments are started, there will undoubtedly be findings that will modify the initial design approach. For this reason, the initial monolithic rectenna development should be started as an exploratory venture. Total rectenna areas and associated power levels should be kept low to minimize the cost of test equipment, particularly the microwave generator which initially would be some other device than the gyrotron which is much too massive and expensive to use for laboratory work.

4.0 SUMMARY OF RESULTS

The scope of the activities carried out under this contract covered two activities whose objectives were distinctly different. The first of these objectives was the establishment of a thin-film, etched-circuit rectenna at 2.45 GHz that would be satisfactory for space operation. The second objective was investigating and establishing, if possible, suitable technologies for constructing rectennas at very high frequencies of 20 GHz and above.

A summary of the results of the work effort is logically divided according to these two different objectives.

4.1 Summary of Results to Improve the Thin-Film, Etched-Circuit Format of the Rectenna and to Adapt it to Space Use

1. In response to a need to determine why the early approach to a thin-film rectenna employing mylar was not satisfactory, special measurement tools were developed and used to determine precisely the dielectric constants and loss tangents of the mylar material and the loss properties of the adhesive material used to bond the mylar to the copper microwave circuits.

2. These measurement tools indicated that the published data on the dielectric constant and loss tangent for mylar were in error by a considerable factor and that our measurements were consistent with the poor behavior of the rectenna element. In the course of investigating the source for this published error, it was found that the correct values for mylar had been subsequently published in a relatively obscure report, thus corroborating our findings.

3. In the course of investigating more suitable materials for the dielectric film it was also found that Kapton had much better characteristics than those that had been published. This finding was based upon our experimental measurements and other information sources. It was decided to redesign the rectenna based upon the use of Kapton.

4. Experiments in bonding Kapton to copper were carried out to create the needed laminate material. Kapton does not bond directly to copper, so the industry has created a film material that uses Kapton as the core material with Teflon sealed to it. The teflon does bond to copper at suitably high temperatures. Our experiments determined that a one mil Kapton core with one half mil of teflon on both sides was a suitable film material.

5. The laminated material with copper on both faces was not available commercially and it was necessary to make the laminated material using special facilities at one of the Raytheon facilities.

6. Using the special laminated material, a rectenna element design was evolved and suitable masks for the etching process were made. The art work involved three successive reiterations, and took into account a rather drastic change in the design of the Schottky barrier diode.

7. A sequence of diode fabrication considerations as well as a trend toward a higher power density rectenna led to a change in packaging the diode from a glass to a ceramic pill package, and in increasing the size of the diode junction by a factor of three.

8. Measurements on the final design of the rectenna element that were made in a closed system, where accurate measurements of efficiency could be made, gave values of 85% with an estimated probable error of $\pm$1.5%. These efficiencies are only 5% less than the best achieved with the rigid, much heavier, and much more costly bar type rectenna construction. This 85% efficiency contrasts with the 70% efficiency of the mylar based element. Because any inefficiency in a space rectenna results in heat that has to be radiated to space, the improved design generates less than half as much heat for a given DC power output of the rectenna.

9. Based upon the design of the individual rectenna element, masks were made for rectenna sections containing 25 elements. Diodes were bonded to these sections. These sections were then mounted on a special fixture for evaluating their performance when immersed in an incident microwave beam of sufficiently large cross section to provide nearly uniform illumination of the rectenna section.

10. Nearly 120 watts (five watts per diode) were obtained from these sections at an estimated efficiency of 85%. The diode temperature rise for this power level was only 50°C.

11. For these tests, a new procedure utilizing the fluoroptic thermometer was evolved to determine the temperature of the diode and other parts of the rectenna element while it was being subjected to various levels of convective air cooling and incident microwave power. The fluoroptic thermometer permits non-invasive measurements of diode and circuit temperatures because it does not interact with the microwave field in any manner and because its thermal capacity and thermal conduction are negligible.

12. A very important achievement of the program was the introduction of a method of testing the efficiency and power handling capability of the rectenna element without resorting to the measurement of the microwave power input. This achievement arose from the fact that the fluoroptic thermometer makes it possible to determine the temperature of the diode, the critical factor in determining an upper limit on the elements' power handling capability, as a function of velocity of cooling air flowing over the surface of the balance of the rectenna element which serves as a cooling radiator for the diode as well as for heat generated elsewhere. Furthermore, the diode temperature for a given air flow can be observed as a function of the power dissipated within the diode resulting from the injection of carefully measured dc power. But when the diode is behaving as a microwave rectifier, diode inefficiency results in heat being generated in the same spot within the diode. This equivalence of heat sources allows the use of information obtained from observing temperature

rise with injection of DC power to determine power being dissipated within the diode when the rectenna element is absorbing microwave radiation and causing the diode temperature to increase. From this equivalence, the rectenna element efficiency can be closely approximated by the formula:

$$\text{Rectenna Element Efficiency} \approx \text{Diode Efficiency} = \frac{\text{DC Power Out}}{\text{DC Power Out} + \text{Diode Dissipation}}$$

The rectenna element efficiency approximates diode efficiency because the diode generates about 80% of the inefficiency and heat within the complete rectenna element. Further, the non-diode circuit losses are greatest in the region of the diode and tend to also raise the temperature of the rectenna element as measured at the diode.

13. The efficiency of individual rectenna elements measured with this technique closely checked those made in a closed system where the microwave power was carefully measured. The new method has the great advantage of allowing simultaneous measurements of efficiency and diode temperature rise while measurements of convective air flow are also being made under conditions that simulate, for example, those found in the application of the rectenna to a wing of an airplane. Such conditions would be difficult to simulate in a closed system for checking a rectenna.

14. Using this new test procedure, we were able to determine the dissipation power occurring within the diode as a function of the measured temperature rise on the case of the diode and of the air flow velocity at sea level pressure. Then, with a reasonable assumption for the efficiency of the diode and corresponding rectenna element, projections could be made of the DC power output density of the rectenna for acceptable diode temperature rise and convective cooling practices.

15. Under the assumption that it is the mass of air flow, or air velocity times density, that determines the amount of cooling, the results obtained at sea level can be transferred to high altitude where the air density is much less but the velocity of air flow may be much greater, for example, in the application to an airplane wing.

4.2 Summary of Results to Develop a Technology for Constructing Rectennas at Frequencies of 20 GHz and Above

1. An important result, after a thorough effort to adapt the thin-film, etched-circuit rectenna technology to a high frequency rectenna, was the conclusion that it would be very difficult to adapt the technology.

2. However, as part of this effort, the interdigital finger technique was introduced to allow lumped capacitances to be formed on one side of the film only. This technique was later applied to thick dielectric substrates.

3. The tentative move to the use of alumina ceramic as both a substrate for the circuit and as the dielectric filler between the foreplane and reflecting plane. In this conceptual arrangement the microwave circuit, with interdigital fingers for capacitors, would be silk screened onto the ceramic. The diodes were visualized as beam-lead devices that were compression bonded to the silk screened circuits.

4. A final move to a monolithic rectenna that used GaAs as a substrate for the microwave circuits and as a base for the integral Schottky barrier diodes.

5. The design of the diode itself, together with a reasonable estimate of its power handling capability as determined by its efficiency and ability of the substrate to conduct heat to the reflecting plane. A quantitative analysis of heat flow was made.

6. The estimate that the power handling capability in terms of DC power output density should be in the range of 3 to 10 kilowatts of DC power per square meter.

7. The development of an experimental approach to matching the rectenna element, when placed on a ceramic (or semiconductor) substrate, to space. It was possible to make the match by greatly reducing the dimensions of the half-wave dipole. The work was carried out at 2.45 GHz but can be scaled to the very high frequencies of interest.

8. The single dipole on its ceramic substrate, attached to the balance of a thin-film, etched-circuit rectenna element, was evaluated for efficiency in a closed system using waveguide and found to exhibit normal, high efficiency.

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Looks like people are still trying to harvest stray electromagnetic waves and convert these into usable energy. Some of you may remember the huge controversy we generated two years ago with news of RCA's Airnergy, a device that claimed to make electricity from WiFi signals. Most of you agreed this was bunk as there was just not enough power in these signals. Ok. But what about microwaves? Japanese company DenGyo has announced the Microwave Regenerative Converter, which is what they call a "rectenna". This stands for antenna and rectifier, smushed together, and not for anal antenna. The idea is that you'd place the device inside a microwave oven and any energy that doesn't go directly towards heating up your food would be converted back into electricity through this device, up to 100W. Since the water content of your food determines the oven's efficiency, the rectenna would be doing the most converting with old stale bread or some similar dry things.

If you follow the links, you'll be treated to all manner of mathematical formulas. We can't understand them, but we think this could work. Any engineers want to set us straight?

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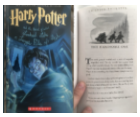
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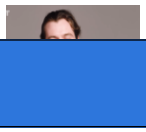
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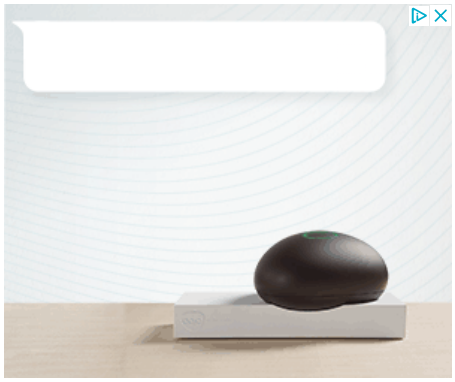

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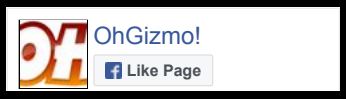

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Development of a Rectenna Adapted to Ultra-wide Load Range for Microwave Power Transmission

Yong Huang

Kyoto University

Doctoral Thesis

Development of a Rectenna Adapted to Ultra-wide Load Range for Microwave Power Transmission

Author:
Yong Huang

Supervisor:
Prof. Naoki Shinohara

A thesis submitted in fulfilment of the requirements for the degree of Doctor of Philosophy

in the

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Abstract

Numerous studies have been carried out on the wireless power transmission (WPT) technologies, which transfers power to the electrical loads without man-made wires. As one mean of WPT, microwave power transmission (MPT) has attracted lots of attention due to its long-distance power transmission capability. A notable problem among these technologies is that the RF to DC conversion efficiency could be easily affected by the load resistance and the input power. Therefore, an impedance matching circuit becomes crucial to realize the wireless power transmission with high efficiency. Moreover, most of the impedance matching circuits in microwave circuits utilize distributed-constant circuits, whereas, the rectenna is a special distributed-constant circuit yielding a DC output. This present thesis aims at developing an ultra-wide load range rectenna for microwave power transmission and a resistance matching circuit from the viewpoint of DC, which is easy to design since it is unnecessary to consider high harmonics.

Different from the conventional DC-DC converters used for voltage conversion, in the present thesis, a resistance conversion DC-DC converter is proposed and validated in the following procedures:

Firstly, with aid of the simulator of advanced design system (ADS), a common pulse-width modulation (PWM) controlled boost converter is checked as an impedance matching circuit for rectifying circuit. For the validation a simulation model of RF-DC-DC circuit is built which consists of a simple single shunt rectifying circuit and the boost converter. The simulation results show that the overall efficiency of the RF-DC-DC circuit is almost constant over 70 % in the load range from 370 to 1300 Ω. Moreover, it is also found that the boost converter can prevent the reverse voltage applied on the rectifying diode from exceeding breakdown voltage and then the rectifying circuit can keep acting at the peak efficiency point. However, this PWM controlled boost converter can only convert

a wide load range into a narrow input resistance range which is insufficient for an impedance matching requirement of rectenna.

Next, according to the previous simulation results, an RF-DC-DC circuit is shown to be useful for impedance matching application. Therefore, an externally-powered RF-DC-DC circuit, consisting of a negative output rectifying circuit and a DC-DC converter, is designed. To choose a more suitable type of DC-DC converter, the input/output resistance relationships of three major DC-DC converter topologies (i.e. buck converter, boost converter and buck-boost converter) are investigated in continuous conduction mode (CCM) and discontinuous conduction mode (DCM), respectively. Only the DCM buck-boost converter is found to exhibit constant input resistance characteristic independent of the input voltage and the load resistance. Therefore, an inverting DCM buck-boost converter is adopted with an extra DC power supply for the control-pulse circuit, and the input resistance of the buck-boost converter is set to be close to the optimal load of the rectifying circuit. The experimental results show that the overall efficiency of this externally-powered RF-DC-DC circuit is approximately constant and over 60 %, despite the load resistance ranging from 100 to 5000 Ω.

However, the externally-powered RF-DC-DC circuit is difficult for the practical applications because of the requirement of an extra DC power supply. To solve this problem a self-powered RF-DC-DC circuit consisting of a positive output rectifying circuit and a non-inverting self-powered buck-boost converter is proposed. It can obtain a positive output voltage with a single positive input voltage. Furthermore, the input resistance of the buck-boost converter is designed to be equal to the optimal load of the rectifying circuit. The experimental results show that the overall efficiency of the self-powered RF-DC-DC circuit is constant and over 66 %, despite an ultra-wide load ranging from 200 to 10000 Ω.

Finally, several experiments are carried out on driving a DC motor using MPT with a compact designed power-receiving device which consists of a rectenna array (including some antennas and rectifiers) and an improved buck-boost converter. There exits four types of power-receiving devices combined by two different rectenna arrays and two different buck-boost converters. With these four power-receiving devices, some experiments are conducted by driving a dynamic load resistance device such as a DC motor with continuous-wave (CW) power transmission and pulsed-wave power transmission. In the CW case, the overall efficiency of the compact power-receiving device is above 50 % in a wide power

density ranging from 0.25 to 2.08 $\mathrm{mW/cm^2}$. In the pulsed-wave case, the overall efficiency is above 44 % in the duty ratio ranging from 0.2 to 1 for a power density of 0.98 $\mathrm{mW/cm^2}$. Moreover, at a fixed duty ratio of 0.5, the overall efficiency is almost constant at 59 % with the pulsed-wave frequency changing from 0.33 to 41.7 kHz.

Therefore, tested by both load resistances and a dynamic load resistance device, the proposed rectenna is found to be excellent in a relatively high efficiency for an ultra-wide load range. As a novel application of DC-DC converter, the proposed buck-boost converter, with constant input resistance characteristic, is also expected to be valuable for other WPT applications.

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Nomenclature

δ	off-state ratio
ε_{r}	dielectric constant
$\tan\,\delta$	dissipation factor
η_{all}	overall efficiency
$\eta_{\mathrm{dc-dc}}$	DC-DC converter efficiency
$\eta_{\mathrm{rf-dc}}$	RF-DC conversion efficiency
η_{prd}	overall efficiency of power-receiving device
η_{track}	tracking efficiency
$\eta_{\mathrm{track}}^{'}$	tracking efficiency
$\eta_{\mathrm{prd}}^{'}$	overall efficiency of power-receiving device
λ	wavelength
ΔI_{Lon}	inductor current variation at on-state
ΔI_{Loff}	inductor current variation at off-state
Δt	time variation
∞	infinity
B_{V}	reverse breakdown voltage
C_{J0}	junction capacitance
C_{in}	input capacitance
C_{o}	output capacitance
D	duty-on ratio
D_{pw}	pulsed-wave duty-on ratio
D_{h}	high frequency duty-on ratio
D_{ι}	low frequency duty-on ratio
dB	decibel
dBi	decibels over isotropic
E_{g}	activation energy
f	frequency
f_{sw}	switching frequency

f_{h}	high frequency
f_{ι}	low frequency
H	dielectric substrate thickness
I_{d}	diode current
I_{dc}	inductor allowable dc current
I_{inave}	converter average input current
I_{in}	converter input current
I_{L}	inductor current
I_{LP}	inductor peak current
I_{m}	DC motor input current
I_{o}	output current
I_{s}	saturation current
I_{SS}	power supply current
I_{BV}	reverse breakdown current
L	inductance
M	junction grading coefficient
N	emission coefficient
P_{RF}	microwave input power
P_{d}	diode loss ratio
P_{rall}	received RF power of antenna array
P_{m}	DC motor input power
P_{mos}	mosfet loss ratio
P_{o}	output power
R_{dc}	inductor dc resistance
$R_{\mathrm{DS(on)}}$	static drain-source on-resistance
R_{L}	load resistance
R_{in}	emulated input resistance of converter
$R_{\mathrm{in}}^{'}$	emulated input resistance of converter
R_{m}	emulated input resistance of DC motor
R_{out}	output load resistance
R_{s}	series resistance of diode
t	copper thickness
T	period
T_{on}	switch-on period
T_{off}	switch-off period
V_{d}	reverse voltage on diode

V_{DC}	dc output voltage
V_{DDmin}	minimum supply input
V_{DSS}	drain-source voltage
V_{FM}	peak forward voltage
V_{gate}	gate control-pulse voltage
V_{GS}	gate-source voltage
V_{h}	high frequency pulse voltage
V_{in}	input voltage
V_{j}	junction potential
V_{L}	inductor voltage
V_{ι}	low frequency pulse voltage
V_{rated}	rating voltage
V_{RRM}	repetitive peak reverse voltage
V_{m}	DC motor input voltage
V_{out}	output voltage
V_{SSmin}	minimum operating supply
Z_{out}	rectifier output impedance
Z_{opt}	rectifier optimal load
Z_{sw}	input impedance of control-pulse circuit

Chapter 1

Introduction

1.1 Wireless Power Transmission

Wireless power transmission (WPT) is the transmission of electrical energy from a power source to an electrical load without man-made conductors, which is useful in cases where interconnecting wires are inconvenient, hazardous, or impossible. Maxwell's equations, which were formulated in 1864, are essentially the first theoretical basis of WPT [1]. The concept of the Poynting vector elucidates the radio wave as representing the energy flux of an electromagnetic field. After Maxwell and Poynting, Nikola Tesla had a dream over a hundred years ago that all electricity would be provided without wires. He carried out the first WPT experiments via 150 kHz radio wave at the end of 19^{th} century [2]. However, his experiments ended in failure because they were far ahead of the high frequency technologies at that time. As the wireless communications and radar remote sensing technologies developing, William C. Brown developed a rectifying antenna named a "rectenna" and restarted WPT experiments with high-efficiency microwave technologies in 1960s [3]. The efficiency of the first rectenna developed by Brown was 50 % at an output of 4 WDC and 40 % at an output of 7 WDC, respectively [4]. With the rectenna, Brown's team had conducted many experiments on WPT such as the successful application of microwave power transmission (MPT) to a wired helicopter in 1964 and a free-flying helicopter in 1968 as shown in Fig. 1.1 [5].

Based on Brown's works, P. E. Glaser proposed a Space Solar Power Satellite (SSPS) system in 1968 [6], which is a huge satellite designed as an electric power

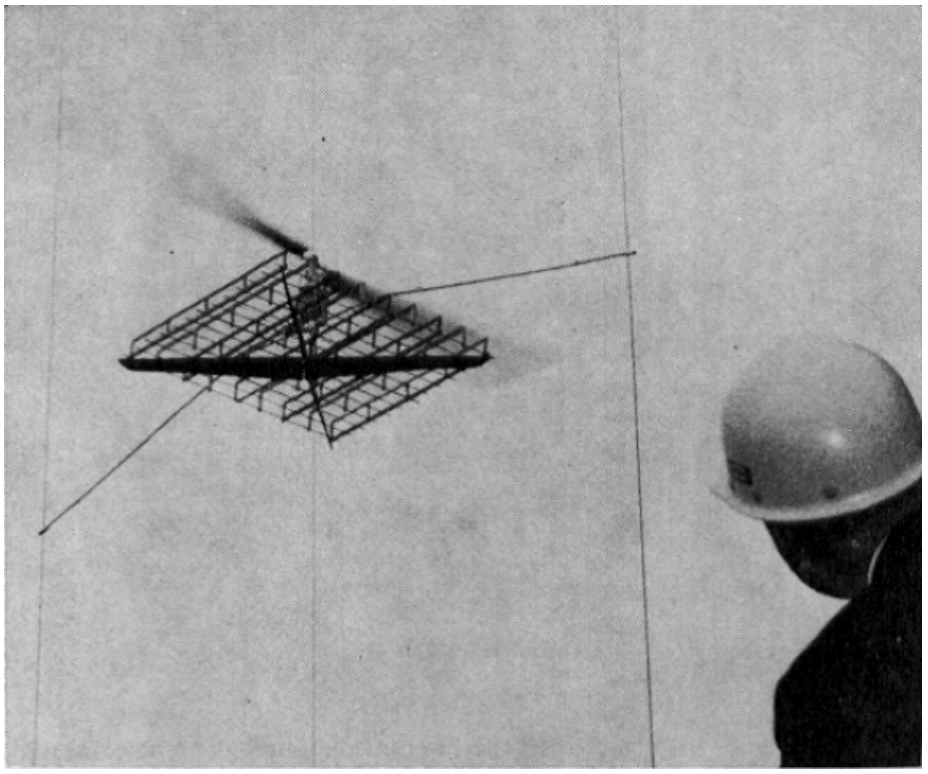


Figure 1.1: MPT experiment by W. C. Brown in 1964: the view of microwave-powered helicopter in flight at an altitude of 50 ft [5].

plant orbiting in the Geostationary Earth Orbit (GEO). Numerous Japanese scientists developed MPT technologies and researches throughout the 1980s [7]-[9]. Hiroshi Matsumoto's team carried out the first MPT experiment in space. The rocket experiment called MINIX (Microwave Ionosphere Nonlinear Interaction eXperiment) was conducted in 1983 as shown in Fig. 1.2, and ISY-METS (International Space Year - Microwave Energy Transmission in Space) was conducted in 1993 and so on.

The historical summary of WPT from Maxwell's times to nowadays is shown in Fig. 1.3 [10]. Over a hundred years development, WPT seems to be difficult for the commercial applications until recent years. A. Kurs's group, from MIT (Massachusetts Institute of Technology) of USA, proposed an efficient WPT via strongly coupled magnetic resonances at 9.9 MHz radio wave which was published on Science in 2007 [11]. They succeeded to transfer 60 watts with the efficiency of 40 % between two coils over distances in excess of 2 meters. Due to their success, it seems that resonant coupling WPT is more suitable for commercial needs. Therefore, there appears a worldwide upsurge of WPT researches in universities and companies.

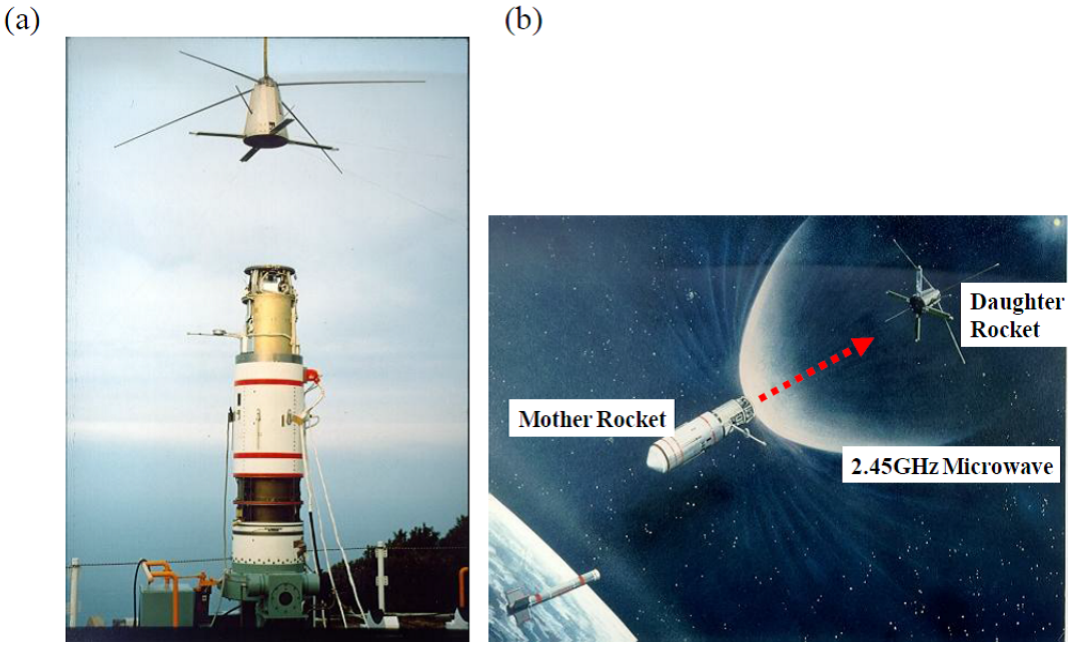


Figure 1.2: The first rocket experiment by H. Matsumoto in Japan in 1983, called the MINIX project. (a) Photo of mother and daughter rockets. (b) Photo of the experiment. These experiments were focused on the nonlinear interaction between intense microwaves and ionospheric plasmas and directed towards SSPS applications [8].

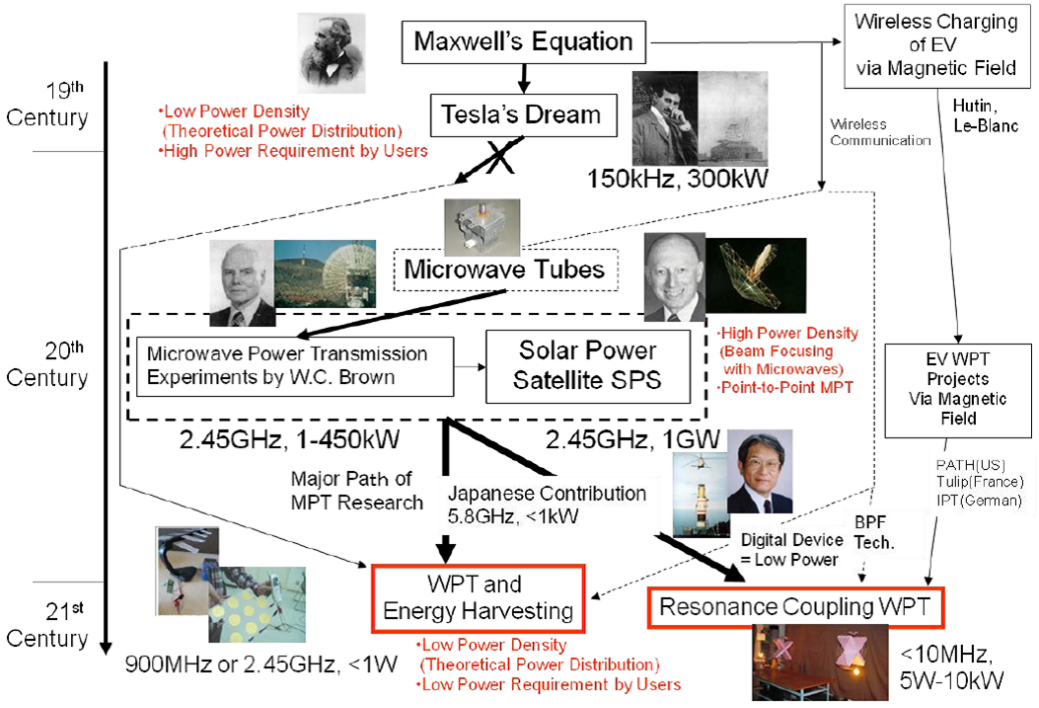


Figure 1.3: Historical summary of WPT from Maxwell's equation to RF harvesting technologies of recent years [10].

Table 1.1: Comparison of three WPT technologies [24].

	Radio wave or microwave	Resonance	Inductive coupling
Field	Electromagnetic (EM)	Resonance (Electric, Magnetic or EM)	Magnetic
Method	Antennas	Resonators	Coils
Efficiency	Low to high	High	High
Distance	Short to long	Medium	Short
Power	Low to high	High	High
Safety	EM	Under discussion (Evanescent)	Magnetic
Regulation	Radio wave or microwave	Under discussion	Under discussion

Nowadays, the common form of WPT is mainly carried out by three ways as follows: by radio waves or microwaves [12]-[15] based on wireless communication technologies and system design; by inductive coupling or be named as inductive power transmission [16]-[19]; by electric, magnetic or electromagnetic resonant coupling [20]-[23]. All of those WPT technologies are based on Maxwell's equations in theory but there are some differences in their applications. The characteristics of those WPT technologies are summarized in Table 1.1 [24].

1.2 Microwave Power Transmission

Microwave Power Transmission (MPT) is one type of WPT that mainly focuses on the frequency of 2.45 GHz, 5.8 GHz of ISM (Industry-Science-Medical) band. Based on the beam control technologies and high power transmission, the biggest advantage of MPT is that it offers the possibility of long-distance power transmission with a high efficiency [25]. Recently, with transmission power from micro-watt

level to kilo-watt level, the MPT technologies have been used in various applications. According to the distance of power transmission, applications of MPT can be roughly classified into space-to-space, space-to-ground and ground-to-ground power transmission. Examples of potential space-to-space applications include MPT for a Mars observation airplane [26] and wireless sensors in the spacecraft [27]. A well-known space-to-ground application is SSPS [28]-[31]. The SSPS consists of mainly three segments: solar energy collectors to convert the solar energy into DC (direct current) electricity; DC-to-microwave converters; and a large antenna array to beam down the microwave power to the ground. The first solar collector can be either photo-voltaic cells or solar thermal turbines. The second segment, DC-to-microwave converters, can be either a microwave tube system or a semiconductor system or the combination of both. The third segment is a gigantic antenna array. The SPS is designed as a huge solar power satellite in geostationary orbit, 36,000 km above the Earth's surface, where there is no cloud cover and no night throughout the year. Since microwave energy is almost not absorbed by air, cloud or rain, SSPS is possible to obtain approximately ten times the solar power than that on the ground. As a stable and CO_2-free green energy source, SSPS is expected to be a main power supply in the future during the late 20^{th} century. However, the development of SSPS has been delayed due to the large cost involved and the need to improve the underlying technologies.

Simultaneously, various ground-to-ground applications have been studied such as MPT for wireless LAN devices [32][33], ZigBee devices [34], sensors in car engine compartment [35], electric vehicles [36], a micro-robot in a pipe [37], small DC motors [38][39], implanted medical devices [40] and so on. Most of those MPT applications involve continuous-wave(CW) power transmission, with the exceptions of ZigBee and wireless LAN devices, which use an intermittent microwave to avoid interference between communication and power transmission [34]. Furthermore, some systems that use only ultra low power can run on energy harvested from ambient RF wave and microwave radiation [41]-[43]. Radio frequency energy exists around us such as TV signals, wireless radio networks, cell phone signals and wireless LAN signals. By using a rectifying circuit linked to a receiving antenna, this free flowing RF energy can be captured and converted into usable DC voltage for electronic applications.

1.3 Objectives

1.3.1 Structure of MPT System

However, because of the low overall efficiency of DC-RF-DC conversion and high cost involved, there are few commercial applications of MPT at present. Figure 1.4 shows a common MPT system, which usually includes a microwave generator, transmitting antennas, receiving antennas, rectifying circuits, a power management and a powered device. Here, an RF power transmission system transmits microwave incident on a receiving antenna (or array). After rectification and management, a DC power can afford to a powered device. In MPT system, the overall efficiency of DC-RF-DC conversion can be separated into three stages. The first stage is the DC-to-RF conversion that refers microwave generation and amplification. The second stage is the RF-to-RF transmission that means the antenna efficiency and the microwave transmission efficiency which are decided by the beam control of the antenna array, the transmission power and the number of antenna elements. The third stage is the RF-to-DC conversion that refers RF rectification and power management. The efficiency of the third stage is decided by the rectifying circuit and power management circuit. This thesis will mainly discuss the third stage.

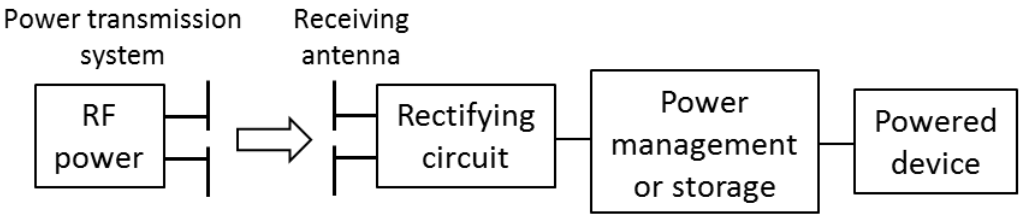


Figure 1.4: Block diagram of a MPT system. An RF power transmission system transmits microwave incident on a receiving antenna (array). After rectification and management, the DC power can afford to power an electrical device.

1.3.2 Overview of Rectenna Studies

The rectifying circuit , as a necessary component of a rectenna, is a very important element for the third stage of MPT system and significantly affects the overall efficiency of the whole MPT system. The single shunt rectifying circuit is the

most common style of rectifying circuit as referred to in Appendix A. The RF-to-DC conversion efficiency of a rectenna depends on the input power and the connected load [44]. As shown in Fig. 1.5, the efficiency of a rectenna is mainly affected by the following three factors. First, the V_{j} (junction potential) effect: The efficiency is low in the low-power region because the voltage applied across the diode is less than or comparable to the forward drop voltage of the diode. Second, the higher harmonics effect: The efficiency increases as the power increases and levels off with the generation of strong higher order harmonics which reduces the proportion of energy that gets converted to DC. Third, the B_{V} (breakdown voltage) effect: The efficiency sharply decreases as the voltage applied across the diode exceeds the breakdown voltage of the diode. It means a rectenna exhibits a maximum efficiency for an optimal output load and lower efficiency for other loads. A maximum efficiency level corresponds to tradeoffs among those three factors of threshold voltage, harmonic generation, and reverse breakdown effects. In most researches on MPT, the load characteristics of the device being powered are matched with the optimal load for the rectenna. Many researches of rectennas at different frequencies have been reported in the literature. This thesis only focuses on the MPT using 2.45 GHz microwave. Table 1.2 lists the overview of relatively high-efficiency rectennas or rectifiers for 2.45 GHz. Until now, the MPT technologies have been developed over fifty years. Most researchers have focused primarily on improving the peak efficiency of rectenna which has reached up to 90 % in recent years [45]. This high efficiency is very close to the efficiency of power transmission by wires. Therefore, it is time to consider a rectenna as a power source for our living.

The efficiency as shown in Table 1.2 is the peak efficiency at a suitable load resistance which is referred as optimal load. However, for the different devices with different load characteristics, it is necessary to use different rectennas with different optimal loads. In other words, the relationship between rectennas and users is 1:1 mode. Furthermore, devices such as a battery, ZigBee sensor nodes and DC motors exhibit variable load characteristics that depend on the operating conditions. These devices will work at very low efficiency or even can not work when they are changed from one operating condition to another because of impedance mismatch. For example, it has been reported that by matching the optimal load of a rectenna array with the load of a DC motor, MPT was successfully used to drive the motor at 3000 rpm, with an efficiency from the antenna array to the motor input port of 50 % [38]. However, if the receiving power of rectenna

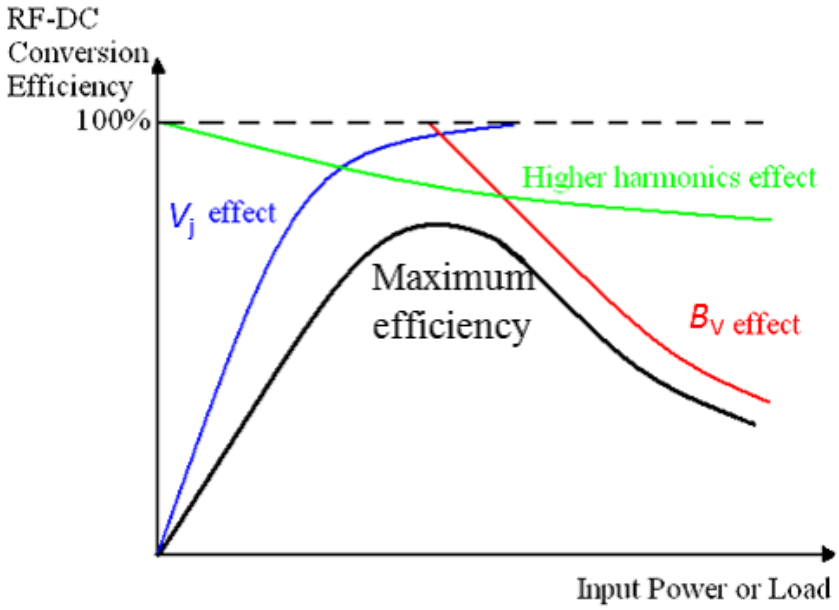


Figure 1.5: Efficiency characteristics of a general rectenna. The efficiency strongly depends on the input power or the load resistance [44].

array is changed because of some obstacles in the microwave transmission route, the efficiency would decrease due to impedance mismatch. Then the rotational speed of the DC motor can not maintain at 3000 rpm. Likewise, an airplane of MILAX (with a rectenna array) droved by a DC motor could not fly as far enough in the experiment [39]. This thesis proposes a flexible rectenna which can exhibit a high efficiency for different load characteristics of users. As shown in Fig. 1.6, the relationship between the flexible rectenna and users is 1:N mode. In this case, the flexible rectenna just seems to be a DC power source for various electronic applications.

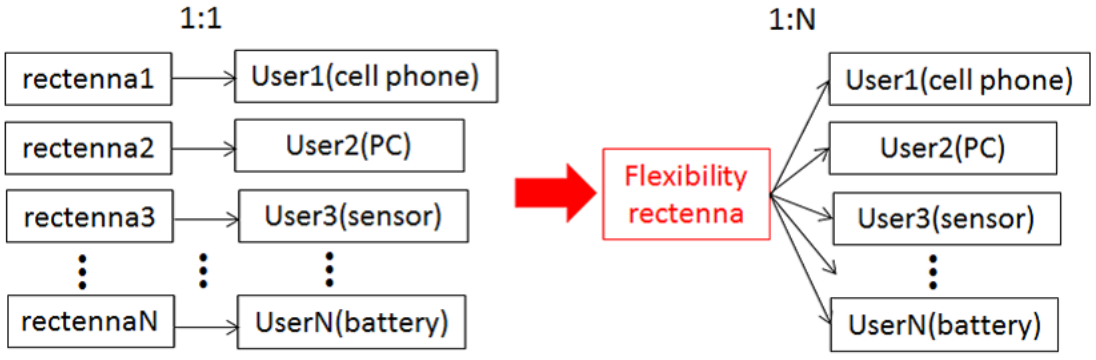


Figure 1.6: Comparison of a general rectenna and a flexible rectenna. A general rectenna for users is 1:1 mode. A flexible rectenna for users is 1:N mode.

In fact, the efficiency depends on the connected load resistance that is also a problem in other forms of WPT systems such as inductive coupling [18] and

Table 1.2: Overview of high-efficiency rectennas or rectifiers for 2.45 GHz microwave with input power levels from mW to W.

Reference	Year	Input power	HF diode	Peak efficiency
[45]	2011	8 W	unknown	90 %
[15]	1996	6.2 W	MA46135-32	81 %
[46]	2008	300 mW	HSMS282c	78 %
[47]	2014	158 mW	HSMS282c	82.3 %
[13]	2002	90 mW	MA4E1317	84.4 %
this work	2013	80 mW	HSMS286F	82.5 %
[48]	2008	32 mW	HSMS2860	75 %
[49]	2010	10 mW	HSMS2860	83 %
[50]	2004	10 mW	HSMS2820	77.8 %

resonance power transmission [22]. The resistance conversion DC-DC converter proposed in this thesis can also be used for impedance matching in other forms of WPT system.

1.4 Thesis Outline

This thesis consists of 6 chapters which mainly focus on the designing and experiment of both rectifying circuits and a resistance conversion DC-DC converter, and the application of MPT for driving a DC motor with a dynamic resistance characteristic. The thesis is organized as follows:

Chapter 1 introduces the development of WPT and the objectives of this thesis.

Chapter 2 utilizes a common pulse-width modulation (PWM) controlled boost converter to play an impedance matching role for a rectifying circuit in the simulation using advanced design system (ADS) simulator. We firstly give an overview of efficiency improvement methods of rectennas and a comparison of impedance matching methods in WPT studies. As a DC-DC converter improves the efficiency of the rectenna from the viewpoint of the DC output, it is easy to design and is independent of the operating frequency. Therefore, a simulation model of an RF-DC-DC circuit which consists of a rectifying circuit and a common boost converter is built. The simulation results indicate that the boost converter can prevent the reverse voltage applied on the rectifying diode from exceeding the breakdown voltage and then the rectifying circuit can keep acting at the peak efficiency point. However, this common boost converter just compresses a wide load range into a narrow load range which is not enough for an impedance matching requirement of the rectenna.

Chapter 3 analyzes the input/output resistance conversion relationships of three basic DC-DC converter topologies in continuous conduction mode (CCM) and discontinuous conduction mode (DCM). Among those six DC-DC converters, the input resistance of the DCM buck-boost converter is independent of the input voltage and the load resistance. Therefore, we design an externally-powered RF-DC-DC circuit which consists of a negative output voltage rectifying circuit and an inverting DCM buck-boost converter. Where the input resistance of the buck-boost converter is close to the optimal load of the rectifying circuit and the buck-boost converter requires an extra DC power source. Finally, the experiment on the externally-powered RF-DC-DC circuit is conducted.

Chapter 4 designs a self-powered RF-DC-DC circuit consisting of a positive output voltage rectifying circuit and a non-inverting DCM buck-boost converter. We firstly design a positive output voltage rectifying circuit with a relatively high peak efficiency. Then the DCM buck-boost converter in Chapter 3 is improved to be a non-inverting self-powered DCM buck-boost converter. Furthermore, the input resistance of the improved buck-boost converter is designed to be equal to the optimal load of the rectifying circuit. Finally, the experiment on the improved buck-boost converter and the self-powered RF-DC-DC circuit are conducted.

Chapter 5 conducts experiments on driving a low-power DC motor with continuous-wave (CW) power transmission and pulsed-wave power transmission. A compact power-receiving device, consisting of a rectenna array and a simplified

buck-boost converter, is used for driving a dynamic resistance characteristic of a DC motor in MPT experiments. Firstly, we design a patch antenna as the power receiving antenna. Then we design two small rectifying circuits with different optimal load resistances. Moreover, two types of improved buck-boost converters, with the control-pulse circuit supplied by the input voltage or the output voltage, are also designed. With the combination of those antennas, two rectifying circuits and two buck-boost converters, there are four combinations of power-receiving devices. Using these four power-receiving devices, MPT experiments on driving a low-power DC motor in both CW and pulsed-wave are conducted.

Finally, Chapter 6 gives a summary and the contributions of this thesis and some directions for future work of MPT area.

Chapter 2

DC-DC Converter for Impedance Matching Application

2.1 Introduction

This chapter firstly gives an overview of efficiency improvement methods and summarizes the impedance matching applications in WPT researches. As a DC-DC converter deals with the DC output of the rectenna, it is easy to design and it plays a resistance conversion role independent of the operating frequency of the rectenna. Therefore, a common PWM controlled boost converter used for impedance matching application in rectifying circuit is investigated. We build a simulation model of an RF-DC-DC circuit which consists of a single shunt rectifying circuit and a PWM controlled boost converter using the simulator of ADS. Through the harmonic balance (HB) analyzing in ADS, the overall efficiency of the RF-DC-DC circuit can keep constant in a wide load range. More importantly, the reason for the constant overall efficiency of the RF-DC-DC circuit is clearly explained.

2.2 Overview of Impedance Matching Methods

The output of the receiving side of WPT is not directly suited to a power supply for circuits because of variations in power and voltage over the load resistance. Accordingly, a power management circuit, which should be able to adapt its input to

the receiving side of WPT and its output to the load, is usually required. Therefore, in order to obtain a necessary DC voltage with high conversion efficiency, there are mainly three solutions reported in the literature. The first method is an active control method which focuses on the active element of the rectifier such as diodes and MOSFET in applications of the RF energy harvesting with an ultra-low RF power input. For example, Gate controlled diodes (GCD) which can achieve the near zero turn-on voltage [51] and Self-V_{th}-Cancellation (SVC) CMOS where the threshold voltage of the MOSFET is cancelled by applying gate bias voltage generated from the output voltage of the rectifier itself [52]. This method intends to improve the V_j effect as shown in Fig. 1.5. The second method proposes a resistance compression networks (RCNs) in the rectifying circuit to reduce the sensitivity of the rectifying circuit versus variations of loads or input powers [53]-[56]. This method can compress a wide load range into a narrow load range in some degree, and the RCNs heavily depends on the operating frequency of the rectifying circuit. The third method uses a DC-DC converter connected to the output of the rectifying circuit for resistance conversion [22][23][43], which focuses on improving the B_V effect as shown in Fig. 1.5. The previous two methods implement solutions at the microwave circuit itself so that their effects are limited by the operating frequency of the rectenna. Their biggest advantage is that the overall efficiency almost do not decrease. The third method adds a DC-DC converter to the DC output of the rectifying circuit which implements solutions on the DC circuit so that it is independent of the operating frequency of the rectifying circuit but the overall efficiency would decrease because of the loss of the DC-DC converter.

Table 2.1 gives a comparison among examples of DC-DC converters, RCNs used in WPT system. Reference [22] used a common buck converter in continuous conduction mode (CCM) for impedance matching in magnetic resonant coupling system. The resistance relationships between the input side and the load resistance of three DC-DC converter topologies, such as buck converter, boost converter and buck-boost converter, were discussed. However, the summarized resistance relationships in [22] did not divide into the continuous conduction mode case and the discontinuous conduction mode (DCM) case. Similarly, in magnetic resonant coupling system, reference [23] utilized a microcontroller in the boost-buck converter to obtain a constant input resistance converter. Therefore, the system of [23] could maintain a constant efficiency in a wider load range than the one of [22]. In [43], it presented a microcontroller-based boost converter with online power stage efficiency optimization and maximum power point tracking in sub-milliwatt levels of

Table 2.1: Comparison of impedance matching methods used in WPT system.

Ref.	Application	Method	Eff. vs. P_{RF}	Eff. vs. R_L
[22]	resonant 13.56 MHz	buck converter in CCM	no discussion	constant from 4.7 to 30 Ω (1:7)
[23]	resonant 13.56 MHz	boost-buck converter	no discussion	constant from 100 to 1000 Ω (1:10)
[43]	RF harvesting 1.96 GHz	boost converter in CRM, DCM	constant from 0.2 to 1 mW (1:5)	no discussion
[56]	RF harvesting 915 MHz, 2.45 GHz	RCNs	low sensitivity	constant from 200 to 1000 Ω (1:5)
[57]	MPT 1.94 GHz, 2.45 GHz	no impedance matching	heavily depend	heavily depend
this work	MPT 2.45 GHz	buck-boost converter in DCM	constant from 30 to 140 mW (1:4.7)	constant from 0.2 to 10 kΩ (1:50)

RF power transmission. It also realized an efficient application of harvesting RF power from a nearby cellular tower with the input power range from 10 μW to 1 mW. In [56], the design of the rectifier was based on the concept of dual-band RCNs as the impedance matching network of the circuit. The 915 MHz/2.45 GHz rectifiers exhibit a low sensitivity at different loads and input powers. Reference [57] summarized some rectennas in different incident power levels and in different frequency, where the rectenna obtained a peak efficiency of 57 % with the power density of 25-200 μW/cm^2 at 2.45 GHz. However, the efficiency was heavily dependent on the incident power and the load resistance since the rectenna did not use any impedance matching circuit. This work proposes an RF-DC-DC circuit consisting of a rectifying circuit and a DCM buck-boost converter. The overall efficiency of the RF-DC-DC circuit exhibits independent of the load and a low sensitivity at different input power variations, which is far superior than other studies. A rectifying circuit connected to a DC-DC converter is firstly referred to

as an RF-DC-DC circuit in this work.

According to the examples reported in Table 2.1, it includes distributed-constant circuits and lumped-constant circuits according to different operating frequencies. Magnetic resonant coupling circuits and DC-DC converters are lumped-constant circuits. Rectifying circuits and RCNs are distributed-constant circuits. Therefore, the impedance matching methods reported in Table 2.1 can be segregated into following three major divisions. First, the main circuit is a distributed-constant circuit and the impedance matching circuit is also a distributed-constant circuit such as references [53]-[56]. In this method, the impedance matching circuit heavily depends on the operating frequency used in the main circuit. Second, the main circuit is a lumped-constant circuit and the impedance matching circuit is also a lumped-constant circuit such as references [22][23]. Third, the main circuit is a distributed-constant circuit while the impedance matching circuit is a lumped-constant circuit such as references [43] and this work. In the later two methods, the impedance matching circuit is a lumped-constant circuit which is independent of the operating frequency used in the main circuit.

2.3 Simulation of Rectifying Circuit

The rectenna is very important for converting RF power into DC power. A rectenna usually contains a receiving antenna, a low pass filter, a rectifying element, a DC pass filter and a load resistance [13][44]. The receiving antenna collects microwave incident power, and the low pass filter matches the antenna impedance to the impedance of the rectifying circuit and rejects higher order harmonics from radiating through the antenna. A high frequency diode placed in shunt across the transmission line is commonly used as a rectifying element. The DC pass filter shorts the RF energy and passes the DC power. Two topologies of DC pass filters, such as a λ /4 line plus a capacitor and a $\lambda/4$ line plus some open stubs, are discussed and compared in [58][59].

This section simulates a simple quarter-wave single shunt rectifier using Advanced Design System (ADS) software of Agilent. Figure 2.1 shows the simulated schematic circuit of the rectifier. TL2, TL3, TL4 are matching circuits. A 100 pF capacitor (C2) is used as a DC block which passes the microwave from input and blocks the DC from the output. The DC pass filter consists of a $\lambda/4$ line (TL1) and

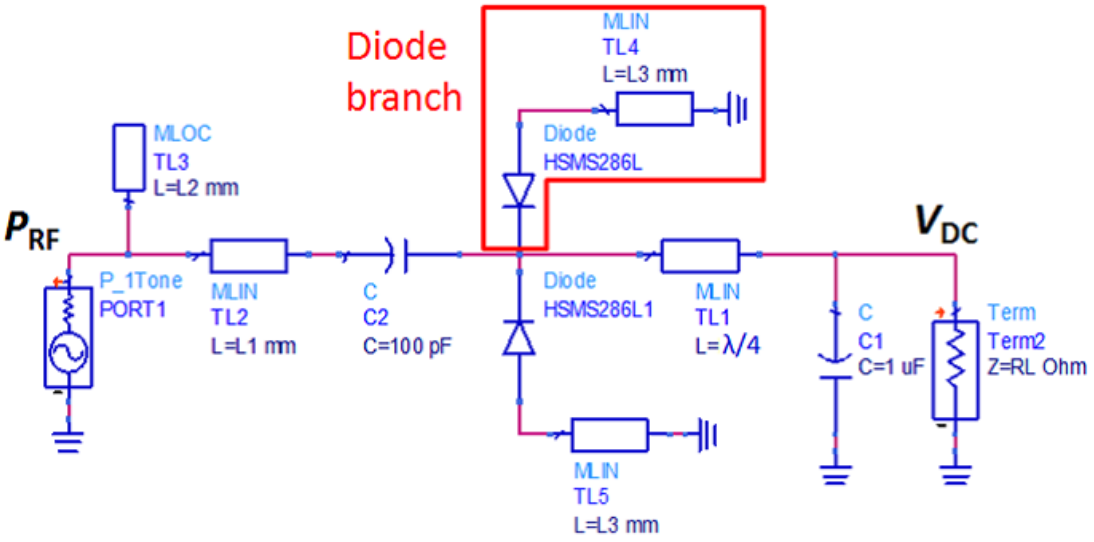


Figure 2.1: Circuit diagram of a simple quarter-wave single shunt rectifier.

Table 2.2: SPICE parameters of the HSMS2860 diode.

Parameter	Value	Unit
B_{V}	7	V
C_{j0}	0.18	pF
E_{g}	0.69	eV
I_{BV}	1E-5	A
I_{s}	5E-8	A
N	1.08	
R_{s}	6	Ω
P_{B}(VJ)	0.65	V
P_{T}(XTI)	2	
M	0.5	

a 1 μF capacitor (C1) which smoothes the DC voltage and reutilizes harmonics energy. The characteristic impedance of all lines is 50 Ω and the fundamental frequency of the input microwave is 2.45 GHz. The two symmetrical diode branches are inserted in the circuit. The RF-DC conversion efficiency is defined only by the diode rectifying efficiency excluding the reflection rate in some literature. Here, the RF-DC conversion efficiency $\eta_{\mathrm{rf-dc}}$ is defined as following, which includes the

reflection.

$$\eta_{\mathrm{rf-dc}} = \frac{V_{\mathrm{DC}}^2}{P_{\mathrm{RF}} R_{\mathrm{L}}} \times 100\,\%, \tag{2.1}$$

where P_{RF}, V_{DC}, R_{L}, as shown in Fig. 2.1, are the input RF power, the DC output voltage and the load resistance, respectively.

As stated in Chapter 1.3.2, the RF-DC conversion efficiency mainly depends on the rectifying diodes. Therefore, it is important to select a suitable rectifying diode. Three diodes in different packages such as HSMS2860, HSMS2865 and HSMS286L of Avago HSMS286x series diode [60] are compared in the simulation. Table 2.2 shows the SPICE parameters of HSMS2860 and Fig. 2.2 shows the packages of three diodes. Using this three diodes in one diode branch or two diode branches, there are total 5 different rectifiers as shown in Table 2.3. These five rectifiers are simulated with the same input power and the comparison of maximum efficiency is shown in Fig. 2.3. The red point shows the maximum efficiency of five rectifiers on the matching state at the optimal load. The efficiency is increasing from 1D to 6D because the total diode loss is decreasing. The reason is that lower series resistance of parallel arranged diodes results in decreasing of diode's normal resistance losses. Therefore, the rectifier with two HSMS286L diode branches is selected in the simulation.

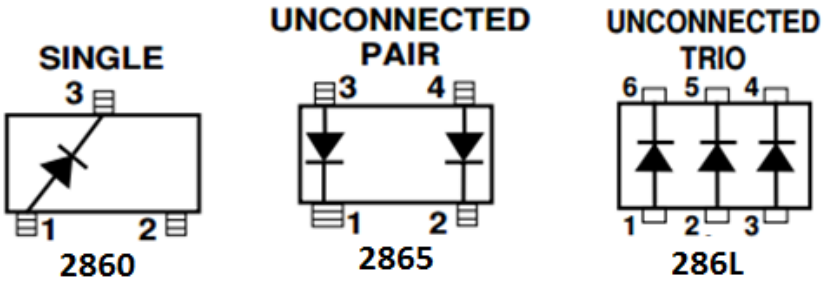


Figure 2.2: Top view of three rectifying diodes.

Then, using Harmonic Balance (HB) analysis of ADS, the rectifying circuit shown in Fig. 2.1 is simulated with the input power of 100 mW. Figure 2.4 shows the simulation results. The maximum RF-DC conversion efficiency is 83.4 % at an optimal load of 95 Ω and the reflection, the diode loss are 0.8 %, 14.7 %, respectively. At the range of load resistance from 50 to 125 Ω, the efficiency is over 70 % but it is strictly decreasing as the load resistance is greater than the optimal load. This is because the diode loss sharply increases as the reverse voltage applied on the diode is over the breakdown voltage. It is clearly seen from the Fig. 2.5 that the maximum voltage applied on the diode is greater than 7 V when the

Table 2.3: Five different simulated rectifiers: combination of three diodes in one diode branch or two diode branches.

Diode No.	Branch No.	Diode
1D	1	2860
2D	1	2865
3D	1	286L
4D	2	2865
6D	2	286L

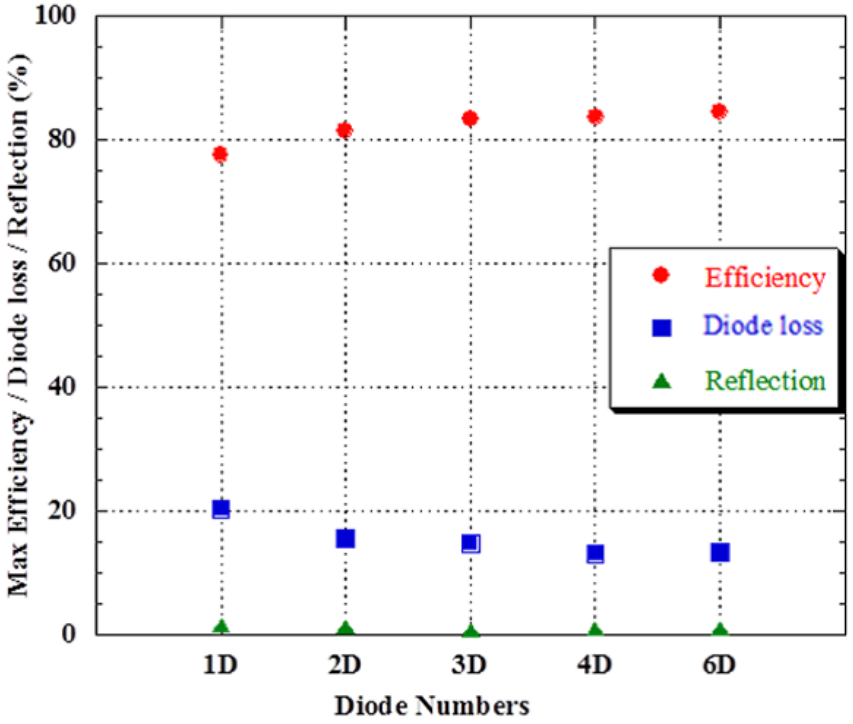


Figure 2.3: Maximum efficiency, diode loss and reflection comparison of five rectifiers.

load resistance is over the optimal load of 95 Ω. The output voltage of rectifier (V_{out}) is 2.8 V at the optimal load of 95 Ω.

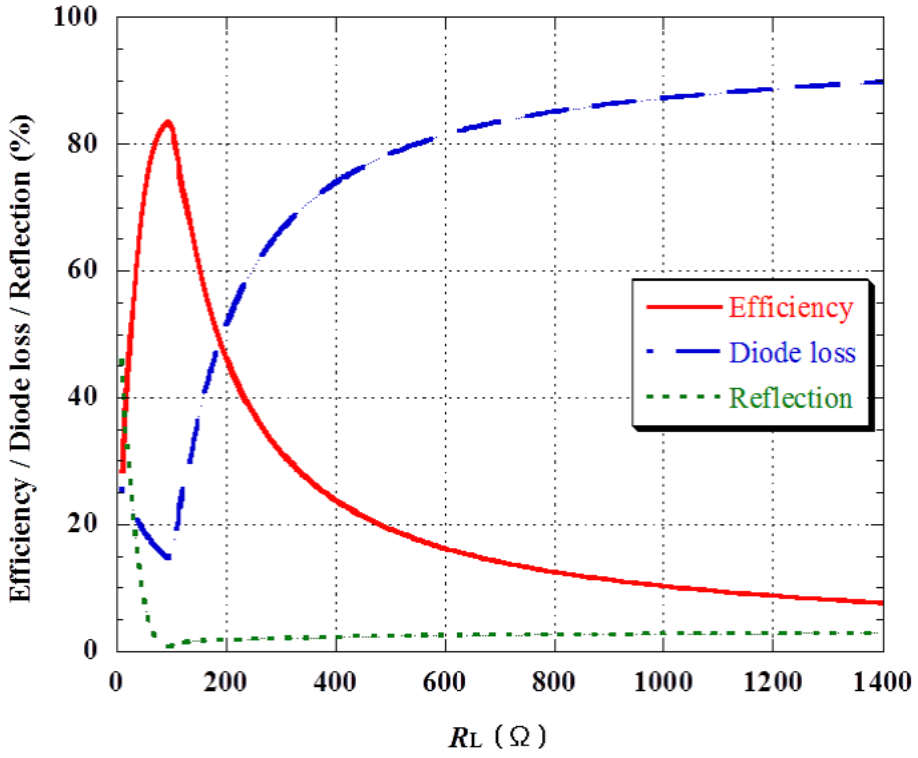


Figure 2.4: Efficiency, diode loss, reflection vs. load resistance of the 6D rectifier (two HSMS286L diode branches).

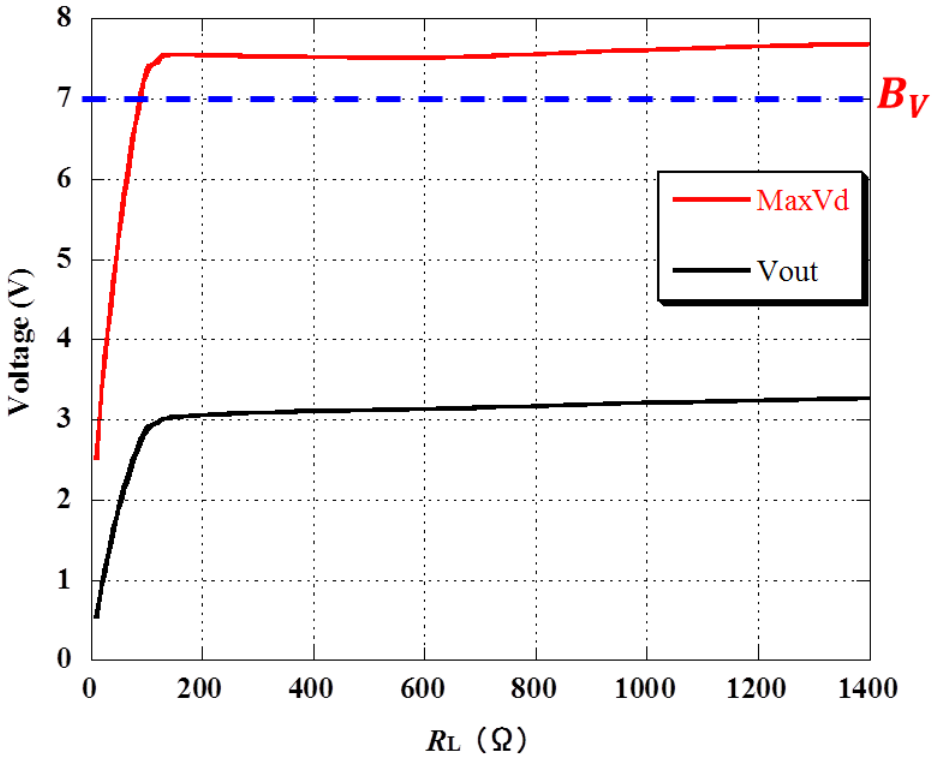


Figure 2.5: Maximum voltage applied on the rectifying diode (MaxV_{d}), output voltage of the rectifier(V_{out}).

2.4 DC-DC Converter Simulation Using ADS

2.4.1 Common DC-DC Converter

Common DC-DC converters are electronic devices usually used whenever we want to change DC electrical power efficiently from one voltage level to another. They are employed in many applications such as computers, office equipments, spacecraft power systems and telecommunications equipments [62]. The basic components of the DC-DC converter can be rearranged to form a buck converter, a boost converter, or a buck-boost converter. The buck converter has a linear control characteristic, and the output voltage is less than or equal to the input voltage. The boost converter is capable of producing an output voltage with a magnitude that is greater than that of the input voltage. The buck-boost converter can either increase or decrease the magnitude of the voltage, but the polarity is inverted. So, with a negative input voltage, the buck-boost converter can produce a positive output voltage of any magnitude.

Most DC converters use MOSFET devices as switching devices because of their very low on-resistance and zero gate-current. When the gate of the MOSFET is off, practically speaking, there is no current, and when the MOSFET is on, there is almost no voltage drop across the drain and the source. Hence, the ideal MOSFET dissipates no power, and the ideal DC-DC converter exhibits 100 % efficiency. In practice, using actual MOSFETs, diodes, and capacitors, DC-DC converter efficiencies of 80 % to 95 % are typically obtained [63]. Additionally, MOSFET probably operates at a high switching frequency which brings about the smaller and lighter transformers, inductors and capacitors. Generally, the converter is used to manage the output voltage with pulse-width modulation (PWM) control [64] or pulse-frequency modulation (PFM) control [65] or PWM/PFM switchable control which uses PFM for low load impedance and automatically switches to PWM control for large load impedance [66].

2.4.2 PWM Controlled Boost Converter

A boost converter is widely applied in energy harvesting systems since the power is very low in an ambient RF environment [61]. In this chapter, a feedback PWM controlled boost converter is used and the schematic circuit is shown in Fig. 2.6.

Table 2.4: Simulation parameters of a PWM controlled boost converter: V_{in} is the input voltage, f_{sw} is the switching frequency of PWM controller.

Parameter	Value	Unit
V_{in}	2.8	V
f_{sw}	1.8	MHz
L	10	μH
C	10	μF
R_1	1	kΩ
R_2	0.5	kΩ

The voltage difference between the feedback node (V_{fb}) and the reference (V_{ref}) is called the error voltage (V_{error}). The PWM wave, generated by comparing the error voltage to the sawtooth wave (V_{saw}), controls the gate of the MOSFET. Table 2.4 shows the simulation parameters. An MBR0520L diode which has a low junction potential voltage of 0.38 V and a high breakdown voltage of 20 V is used. The circuit is simulated by HB analysis of ADS and Fig. 2.7 shows the efficiency, MOSFET loss (P_{mos}), diode loss (P_{d}) versus the load resistance. The efficiency is over 84 % as the load is varying from 370 to 1300 Ω. The maximum efficiency is 89.1 % with the P_{mos} of 5.9 % and the P_{d} of 5 %.

Here, the input resistance of the DC-DC converter (R_{in}) is defined as Eq. (2.2) where the V_{in}, I_{inave} are the input voltage and the average of input current, respectively. Figure 2.8 shows the simulated results of R_{in} which are compressed in the range of 80 to 105 Ω with the load range from 370 to 1300 Ω.

$$R_{\text{in}} = \frac{V_{\text{in}}}{I_{\text{inave}}}. \tag{2.2}$$

2.5 RF-DC-DC Circuit

The RF-DC-DC circuit is a combination of RF-DC rectifier and the boost converter as mentioned above. Fig. 2.9 shows the block diagram of the RF-DC-DC circuit where the boost converter is connected to the RF-DC rectifier. Therefore, the

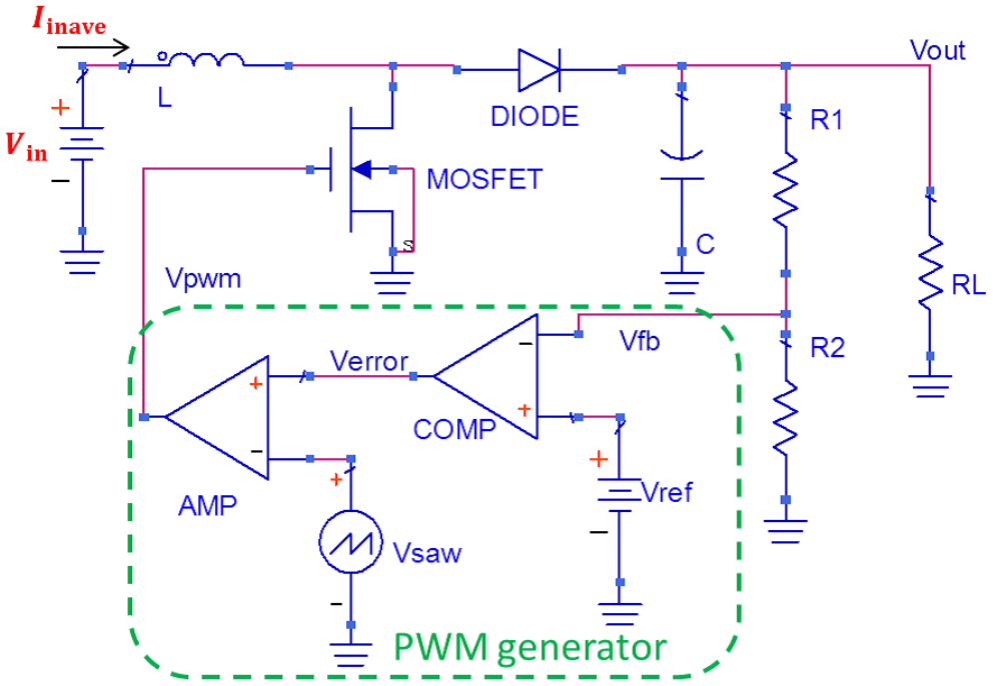


Figure 2.6: Circuit diagram of the simulated DC-DC converter: a PWM controlled boost converter.

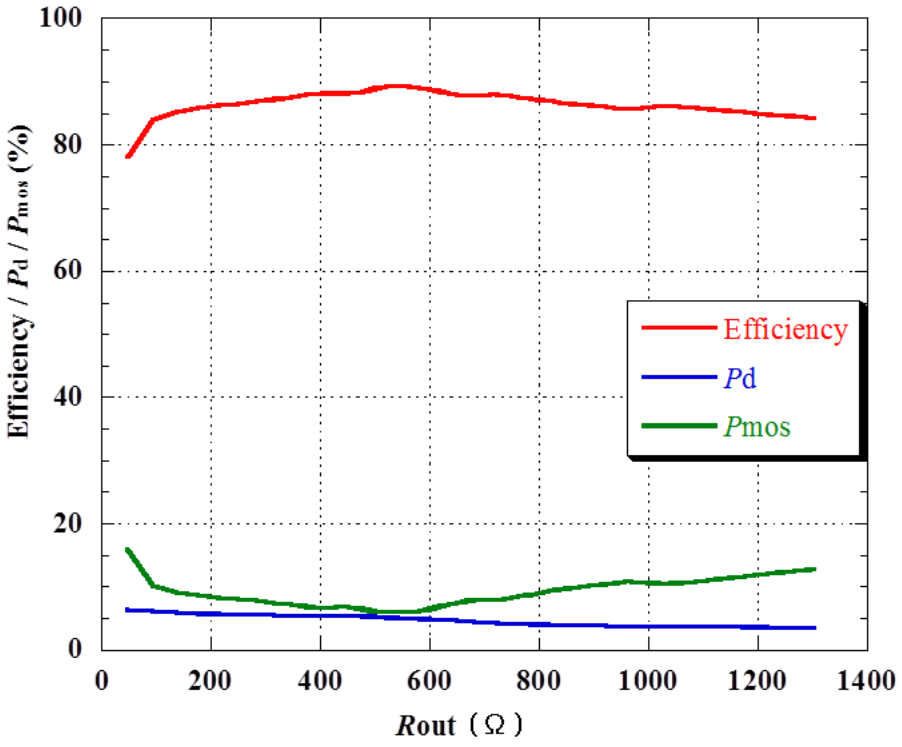


Figure 2.7: Simulation results of the boost converter such as Efficiency, MOSFET loss (P_{mos}), diode loss (P_{d}) vs. load resistance.

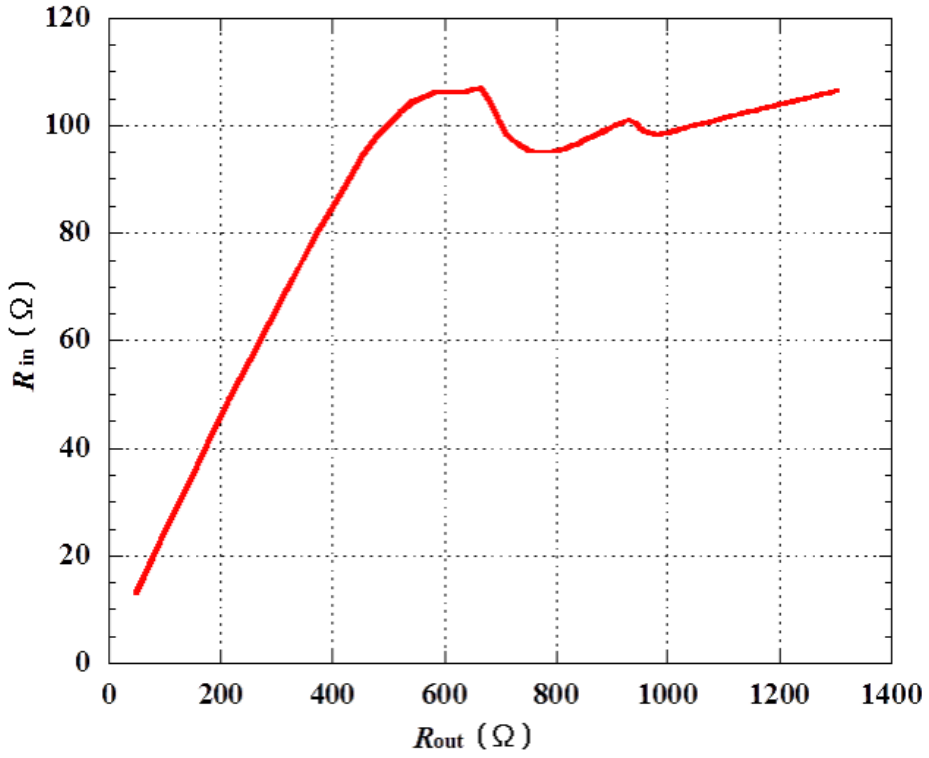


Figure 2.8: Simulation results of the boost converter such as the input resistance (R_{in}) vs. load resistance.

overall efficiency of the RF-DC-DC circuit η_{all} should satisfy following equation.

$$\eta_{\text{all}} = \eta_{\text{rf-dc}} \cdot \eta_{\text{dc-dc}}, \tag{2.3}$$

where $\eta_{\text{dc-dc}}$ is the efficiency of the boost converter.

The relationships between efficiency and load are shown in Table 2.5. As the input resistance R_{in} of the boost converter is matching with the output impedance of rectifier, the overall efficiency of the combined RF-DC-DC circuit is expected to be over 67 %.

Using HB analysis in ADS at two fundamental frequencies of 2.45 GHz and 1.8 MHz to simulate the RF-DC-DC circuit which is shown in Fig. 2.10. The input microwave power is also 100 mW which is the same as the RF-DC rectifier. Figure 2.11 shows the efficiency comparison of the RF-DC-DC circuit and the RF-DC rectifier. Comparing to a narrow range of 50 to 125 Ω of the RF-DC rectifier, the overall efficiency of the RF-DC-DC circuit is almost constant greater than 70 % at the load range of 370 to 1300 Ω. Figure 2.12 shows the impedance comparison which indicates that the Z_{out} of the RF-DC-DC circuit is almost the

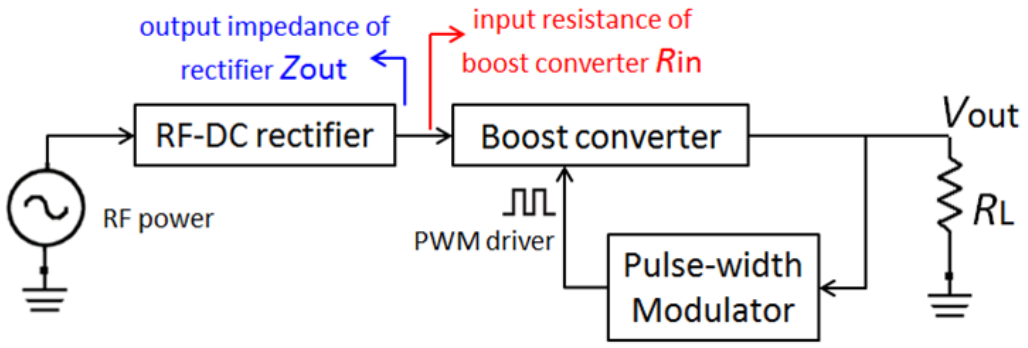


Figure 2.9: Block diagram of the RF-DC-DC circuit: the boost converter is connected to the RF-DC rectifier.

Table 2.5: Efficiency and load relationships of the RF-DC rectifier and boost converter.

	R_{out}	R_{in}	Efficiency
RF-DC rectifier	70-105 Ω		80-83.4 %
boost converter	370-1300 Ω	80-105 Ω	84-89 %
RF-DC-DC circuit	370-1300 Ω		predicted value 67 %

same as the R_{in} of the boost converter. Therefore, the simulation results of the RF-DC-DC circuit agree well with that of the two circuits simulated separately. A comparison of the maximum reverse voltage on the rectifying diode (V_{d}) of the two circuits is shown in Fig. 2.13. It is clear that V_{d} of the RF-DC-DC circuit is constant under 7 V (breakdown voltage) while V_{d} of the RF-DC rectifier is over 7 V when the load resistance is bigger than the optimal load. This results mean that the boost converter limits the maximum reverse voltage of the rectifying diode to prevent the conversion efficiency of the RF-DC rectifier from sharply decreasing.

Finally, the output voltage of the RF-DC-DC circuit is increasing as the load increases which is shown in Fig. 2.14. Here, the output voltage is different with the constant output voltage of the conventional PWM controlled boost converter. According to Figs. 2.13 and 2.14, a constant overall efficiency of the RF-DC-DC circuit can be explained as follows. In the RF-DC-DC circuit, the output voltage of the rectifier (when it is connected with the converter) is almost constant even though the load is changed, which leads to the maximum voltage applied on the

diode being constant and under the breakdown voltage. As a result, the efficiency of the rectifier (when it is connected with the converter) is almost constant which leads to a constant overall efficiency of the RF-DC-DC circuit even though the load is changed. In contrast, the output voltage of the rectifier (when it is not connected with the converter) is increasing as the load is increased, which leads to the maximum voltage applied on the diode increasing up to over the breakdown voltage and the increase of diode losses. Consequently, the efficiency of the rectifier (when it is not connected with the converter) decreases at a high load resistance.

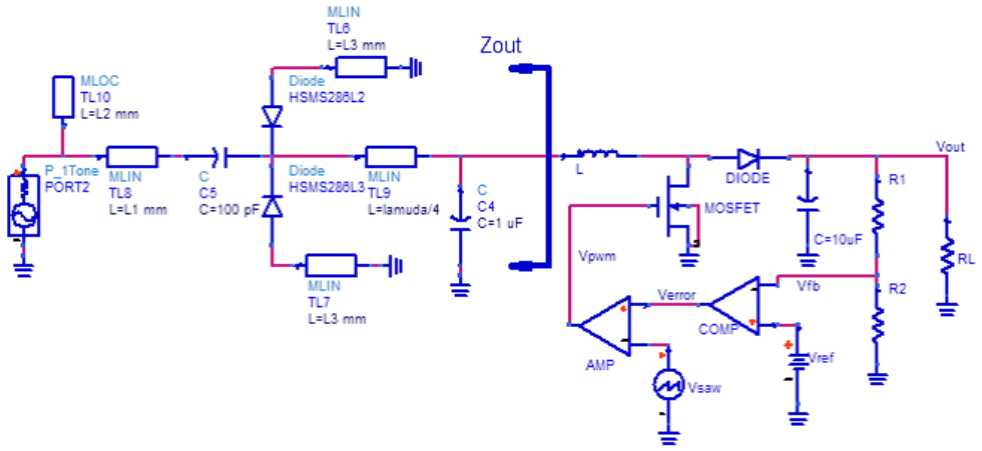


Figure 2.10: Circuit diagram of the RF-DC-DC. Simulate it with two fundamental frequencies of 2.45 GHz and 1.8 MHz using HB analysis in ADS.

2.6 Summary

Firstly, we give an overview of efficiency improvement methods in rectenna researches and a comparison of impedance matching methods in WPT studies. The DC-DC converter addresses the DC output of the rectifying circuit and improves the B_v effect in the efficiency-load characteristics. More importantly, the DC-DC converter as an impedance matching application which is independent of the operating frequency of the rectenna. Therefore, a PWM controlled boost converter is used to improve the efficiency-load characteristics of the rectifying circuit. Using HB analysis of ADS, we build a model of RF-DC-DC circuit which consists of the single shunt rectifying circuit and the PWM controlled boost converter. The RF-DC-DC circuit obtains a steady high overall efficiency (over 70 %) at the load

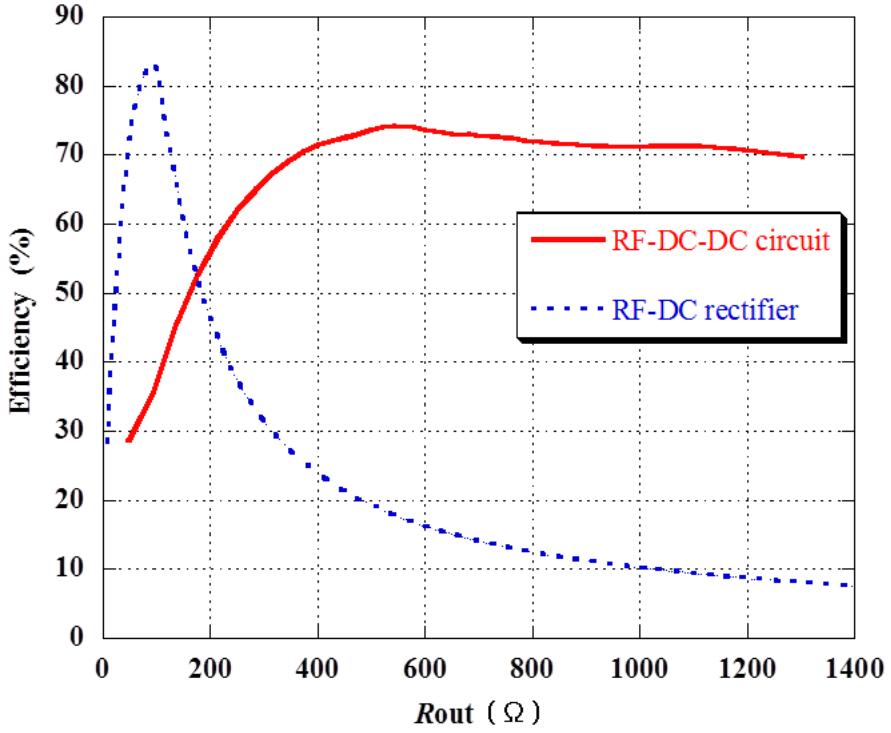


Figure 2.11: Efficiency comparison of the RF-DC-DC circuit and general rectifier.

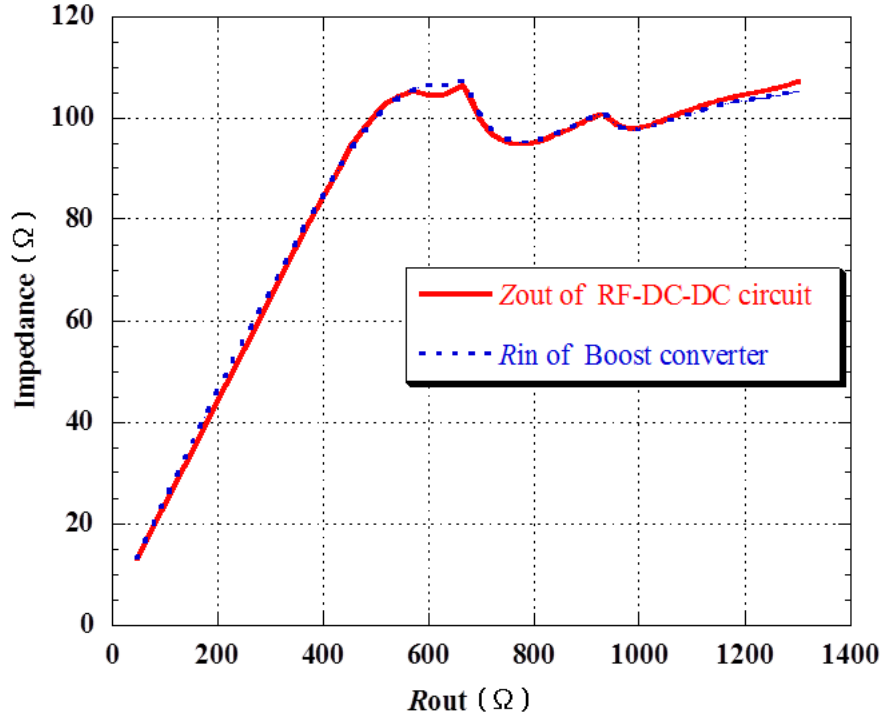


Figure 2.12: Impedance comparison: Z_{out} of the RF-DC-DC circuit is almost the same as R_{in} of the boost converter.

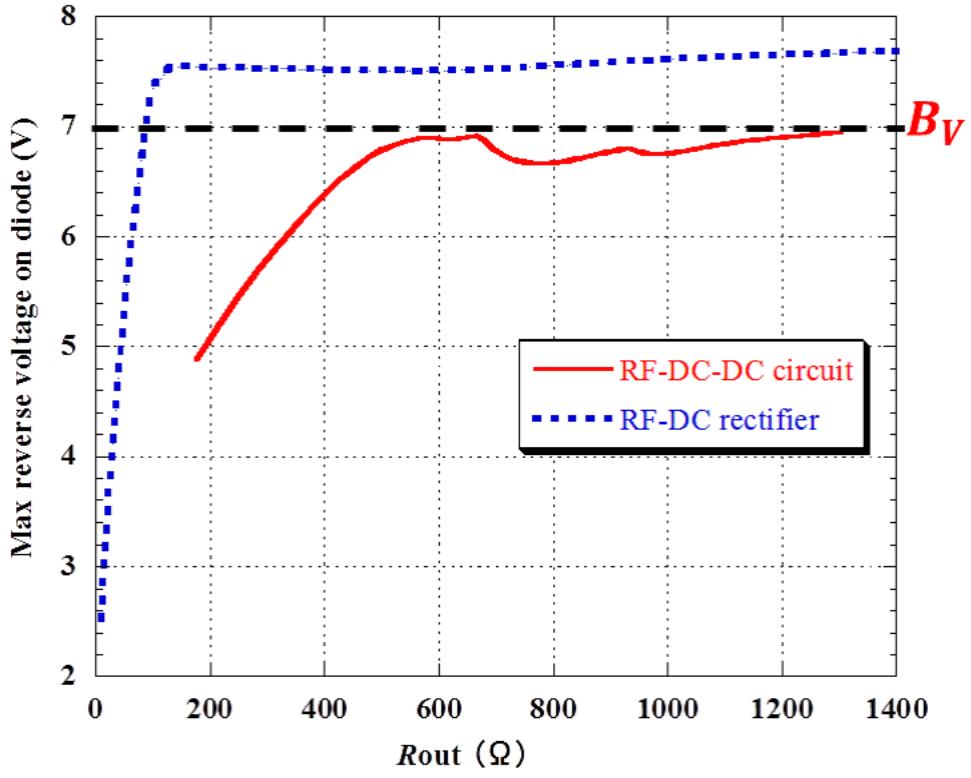


Figure 2.13: Maximum reverse voltage on the rectifying diode (MaxV_d) comparison: MaxV_d of the RF-DC-DC circuit is under breakdown voltage (7 V).

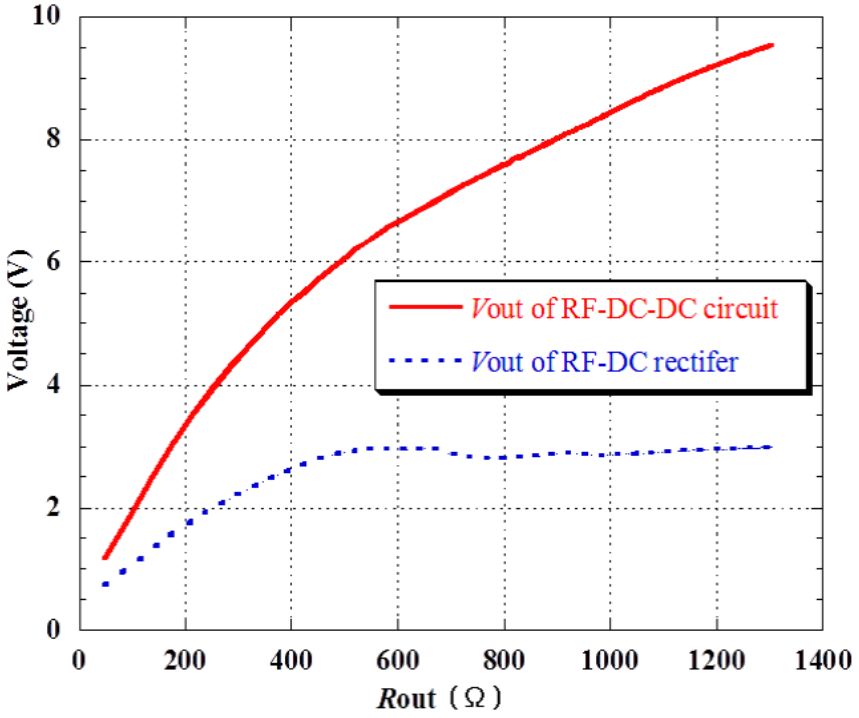


Figure 2.14: Output voltage of the RF-DC-DC circuit and the voltage at the output port of RF-DC rectifier.

range of 370 to 1300 Ω, where the load range is wider than the RF-DC rectifier at the same efficiency level. Additionally, that the boost converter is connected to the RF-DC rectifier can limit the maximum reverse voltage applied on the rectifying diode under breakdown voltage at this load range. The simulation results clearly explain the reason why a DC-DC converter has an effect on impedance matching for RF-DC rectifier. Therefore, an RF-DC-DC circuit is verified to be available for improving efficiency-load relationships of the rectenna in this chapter. However, the simulated PWM controlled boost converter just compresses a wide output load range into a narrow input resistance range in some degree. It is insufficiency for impedance matching for a flexible rectenna. Therefore, based on the simulation results of this chapter, a much more effective DC-DC converter is proposed and discussed in the next chapter.

Chapter 3

Externally-powered RF-DC-DC Circuit

3.1 Introduction

Last chapter demonstrates that a DC-DC converter is useful for impedance matching of a rectifying circuit based on simulation. In this chapter, to select a suitable DC-DC converter for impedance matching of the rectenna, the three basic topologies DC-DC converters such as buck converter, boost converter, buck-boost converter in CCM and DCM are discussed. The CCM is defined by continuous inductor current which is greater than zero over the entire switching period. The DCM is defined by discontinuous inductor current which equals to zero during a portion of the switching period. Hence, the energy stored in the inductor is zero at the beginning and at the end of each switching period. The CCM is preferred for high efficiency and good utilization of semiconductor switches and passive components. The DCM may be used in applications with special control requirements because the dynamic order of the converter is reduced. It is uncommon to mix these two operating modes because of different control algorithms [67].

Both CCM and DCM are discussed for three basic topologies DC-DC converters in this chapter. Unlike the usual discussion on DC-DC converter, this chapter will analysis the relationships between input resistance and load resistance instead of input/output voltages. Comparing those DC-DC converters, an inverting buck-boost converter in DCM, which exhibits constant input resistance,

is selected. The control-pulse circuit of this buck-boost converter requires an extra DC power source. Additionally, to match the polarity of the input voltage and the output voltage, a negative output voltage rectifier is designed. Then, an externally-powered RF-DC-DC circuit consisting of the rectifier and the buck-boost converter is proposed. Finally, experiments on the externally-powered RF-DC-DC converter are conducted.

3.2 DC-DC Converter in CCM and DCM

3.2.1 DC-DC Converter in CCM

A DC-DC converter operating in CCM or DCM depends on the characteristics of the inductor current. In the case of the CCM, the inductor current is continuous and never falls to zero. Figure 3.1 shows the operating waveform of the inductor (I_{L}) in CCM at the steady condition. The I_{LP} is the peak inductor current and I_{LV} is the minimum inductor current which is greater than zero. Figure 3.2 shows the schematic circuits of three basic topologies DC-DC converters which consist of capacitors, an inductor, an N-channel MOSFET (N-MOS), and a diode. T_{on}, T_{off}, T are the on-state time, off-state time and the period, respectively. D is the duty-on ratio of the control-pulse. Then it yields:

$$D = \frac{T_{\mathrm{on}}}{T} = \frac{T_{\mathrm{on}}}{T_{\mathrm{on}} + T_{\mathrm{off}}}. \tag{3.1}$$

As we know, the relationship between the inductance L, the inductor voltage (V_{L}) and the inductor current (I_{L}) satisfies the following formula:

$$V_{\mathrm{L}} = L\frac{\Delta I_{\mathrm{L}}}{\Delta t}. \tag{3.2}$$

Moreover, ΔI_{Lon}, ΔI_{Loff} are the current variation of the inductor at on-state and off-state, respectively. As the overall current variation of the inductor should be zero when the converter operates at steady-state. Then it means:

$$\Delta I_{\mathrm{Lon}} + \Delta I_{\mathrm{Loff}} = 0. \tag{3.3}$$

A). Buck converter

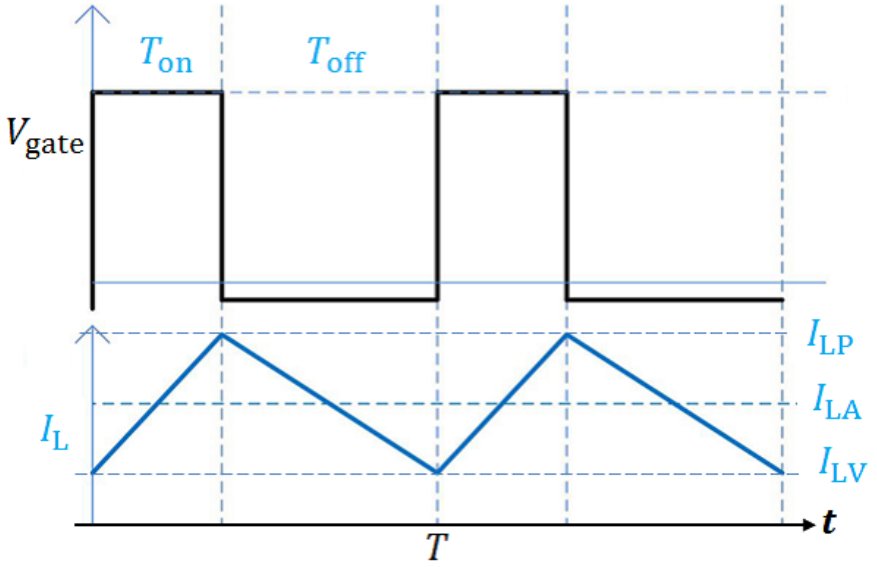


Figure 3.1: Operating waveforms of the inductor current (I_L) and gate control voltage (V_{gate}) for a DC-DC converter in CCM.

As can be seen from the Fig. 3.2(a), when the gate voltage V_{gate} is at high level, N-MOS lets the current flow through drain to source then the current flows through the inductor to the load. At this on-state, the inductor current is increasing to the peak value I_{LP} and no current flows through the diode since it is reverse-biased by the voltage. Then the inductor voltage at on-state (V_{Lon}) can be written as:

$$V_{Lon} = V_{in} - V_o. \tag{3.4}$$

Using Eqs. (3.2) and (3.4), the current variation of the inductor ΔI_{Lon} can be calculated as:

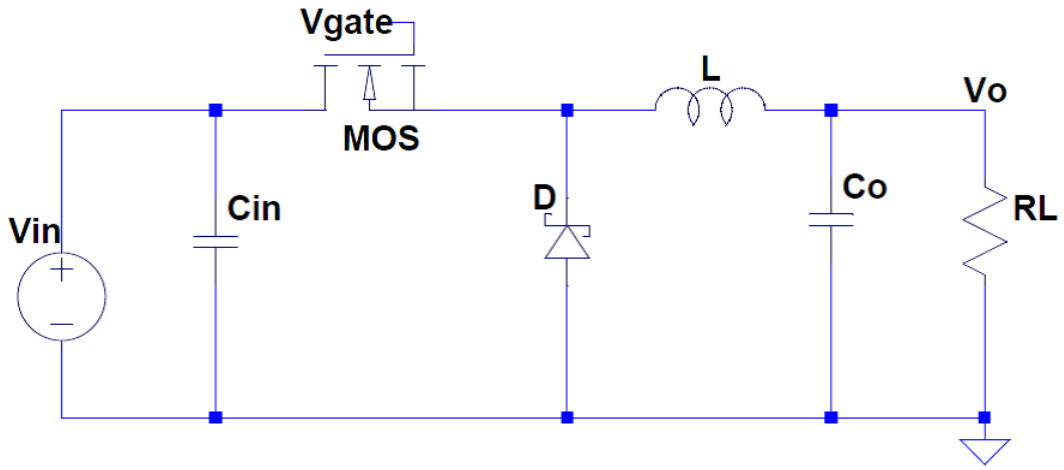
$$\Delta I_{Lon} = \frac{V_{in} - V_o}{L} DT. \tag{3.5}$$

When the gate voltage V_{gate} is at low level (or zero), N-MOS does not let the current flow through drain to source, then the current stored in the inductor flows through the load and diode. At this off-state, the inductor current is decreasing to the valley value I_{LV} and the current flows through diode as it is forward-biased. Then the inductor voltage at off-state (V_{Loff}) can be written as:

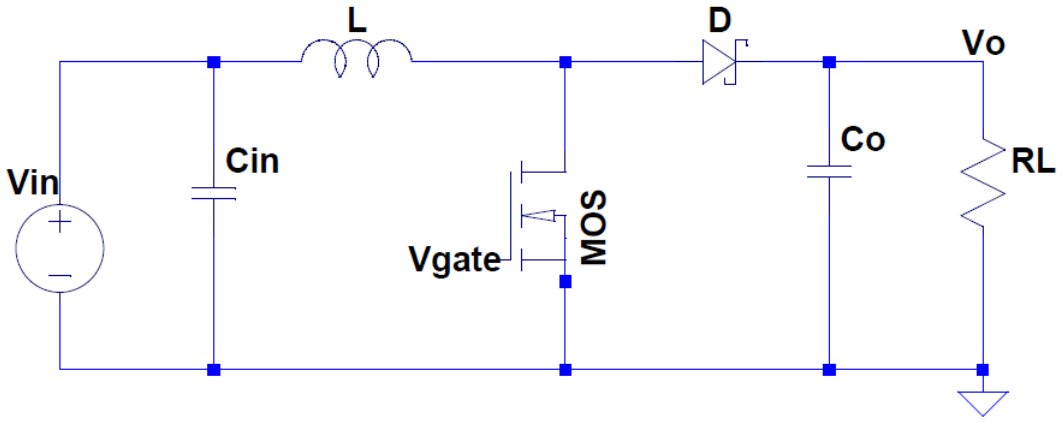
$$V_{Loff} = -V_o. \tag{3.6}$$

Similarly, using Eqs. (3.2) and (3.6), the current variation of the inductor ΔI_{Loff} can be calculated as:

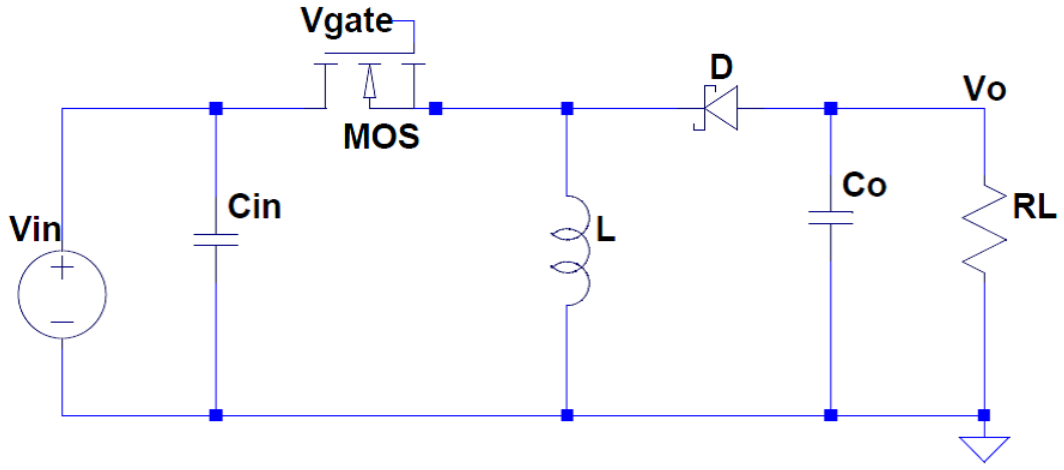
$$\Delta I_{Loff} = -\frac{V_o}{L}(1 - D)T. \tag{3.7}$$



(a)



(b)



(c)

Figure 3.2: Circuit diagram of three basic topologies DC-DC converters. (a) Buck converter. (b) Boost converter. (c) Buck-boost converter.

Based on Eqs. (3.3), (3.5) and (3.7), it yields:

$$V_{\mathrm{o}} = DV_{\mathrm{in}}. \tag{3.8}$$

R_{in} is the average input resistance of the converter and R_{L} is the load resistance. P_{in}, P_{o} are the input power and output power, respectively. Then:

$$P_{\mathrm{in}} = \frac{V_{\mathrm{in}}^2}{R_{\mathrm{in}}}, \tag{3.9}$$

$$P_{\mathrm{o}} = \frac{V_{\mathrm{o}}^2}{R_{\mathrm{L}}}. \tag{3.10}$$

Assuming it is an ideal converter, the input power P_{in} should equal the output power P_{o}. Then it means:

$$\frac{V_{\mathrm{in}}^2}{R_{\mathrm{in}}} = \frac{V_{\mathrm{o}}^2}{R_{\mathrm{L}}}. \tag{3.11}$$

According to Eqs. (3.8) and (3.11), R_{in} can be written as:

$$R_{\mathrm{in}} = \frac{1}{D^2} R_{\mathrm{L}}. \tag{3.12}$$

As $D < 1$, then, $R_{\mathrm{in}} > R_{\mathrm{L}}$ for the buck converter.

B). Boost converter

Similarly, as shown in Fig. 3.2(b), the current flows through the inductor to the ground then the diode and load are shorted at on-state. The inductor current is increasing linearly to the peak value I_{LP}. At off-state, the current flows through the inductor and diode to the load. Then, the inductor current is decreasing to the valley value I_{LV}. This yields following equations.

$$V_{\mathrm{o}} = \frac{1}{1-D} V_{\mathrm{in}}, \tag{3.13}$$

$$R_{\mathrm{in}} = (1-D)^2 R_{\mathrm{L}}. \tag{3.14}$$

Hence, as for boost converter, $R_{\mathrm{in}} < R_{\mathrm{L}}$.

C). Buck-boost converter

As shown in Fig. 3.2(c), the current flows through the inductor to the ground at on-state. The inductor current is increasing linearly to the peak value I_{LP}. At off-state, the current flows through the inductor to the load and diode. Then,

the inductor current is decreasing to the valley value I_{LV}. The polarity of the output voltage V_{o} is negative because the converter inverts the polarity of input and output voltages. This gives following equations.

$$V_{\mathrm{o}} = -\frac{D}{1-D}V_{\mathrm{in}}, \tag{3.15}$$

$$R_{\mathrm{in}} = (\frac{1}{D} - 1)^2 R_{\mathrm{L}}. \tag{3.16}$$

Hence, as for buck-boost converter, $0 < R_{\mathrm{in}} < \infty$. It can be smaller than R_{L} or bigger than R_{L}.

Therefore, we summarize the relationships of input/output resistances and input/output voltages for these three basic topologies DC-DC converters in CCM in Table 3.1.

Table 3.1: Input/output resistance and voltage relationships of three basic topologies DC-DC converters in CCM.

Topology	Voltage	Resistance	Variation of R_{in}
buck	$V_{\mathrm{o}} = DV_{\mathrm{in}}$	$R_{\mathrm{in}} = \frac{1}{D^2}R_{\mathrm{L}}$	$R_{\mathrm{L}} < R_{\mathrm{in}} < \infty$
boost	$V_{\mathrm{o}} = \frac{1}{1-D}V_{\mathrm{in}}$	$R_{\mathrm{in}} = (1-D)^2 R_{\mathrm{L}}$	$0 < R_{\mathrm{in}} < R_{\mathrm{L}}$
buck-boost	$V_{\mathrm{o}} = -\frac{D}{1-D}V_{\mathrm{in}}$	$R_{\mathrm{in}} = (\frac{1}{D} - 1)^2 R_{\mathrm{L}}$	$0 < R_{\mathrm{in}} < \infty$

3.2.2 DC-DC Converter in DCM

In some cases, the amount of energy required by the load is small enough to be transferred in a time not the whole commutation period. In this case, the DC-DC converter operates in DCM, where the current through the inductor falls to zero during a part of the period. Figure 3.3 shows the operating waveform of the inductor (I_{L}) in DCM at the steady condition. There are three states according to three time intervals. 0 - $t1$ is the on-state that the inductor current is increasing linearly; $t1$ - $t2$ is the off-state that the inductor current is decreasing linearly; $t2$ - T is the zero-state that the inductor current is zero. Those time intervals satisfy

following equations:

$$t1 = DT, \tag{3.17}$$

$$t2 - t1 = \delta T. \tag{3.18}$$

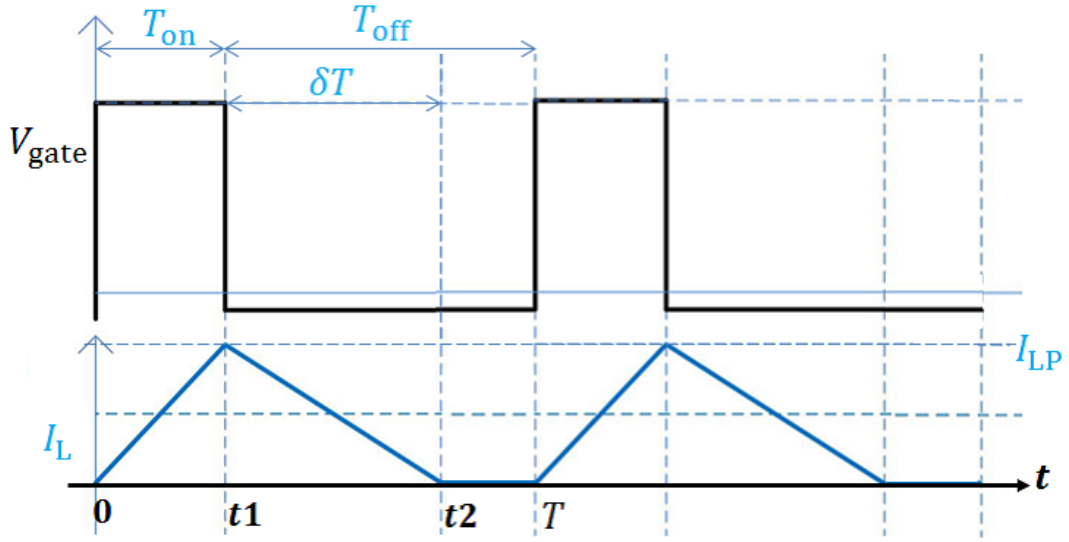


Figure 3.3: Operating waveforms of the inductor current (I_L) and gate control voltage (V_{gate}) for a DC-DC converter in DCM.

The only difference with the CCM is that the inductor is completely discharged at the end of the commutation cycle. Although slight, the difference has a strong effect on the output voltage equation and the input resistance equation.

A). Buck-boost converter

Referring to the Fig. 3.2(c), the current flows through the N-MOS to the inductor at on-state which makes the input voltage (V_{in}) appear across the inductor. The inductor current is increasing from 0 to the peak value I_{LP}. Considering Eqs. (3.2) and (3.17), inductor current variation at on-state ΔI_{Lon} can be written as following:

$$\Delta I_{Lon} = \frac{DT}{L} V_{in}. \tag{3.19}$$

At the off-state, the current flows through the inductor to the load and diode until the inductor current falls to zero through δT period. Assuming zero voltage drop across the diode, the output voltage (V_o) appears across the inductor. Then the inductor current variation at off-state ΔI_{Lon} is given by:

$$\Delta I_{Loff} = \frac{\delta T}{L} V_o. \tag{3.20}$$

Applying Eqs. (3.19) and (3.20) to Eq. (3.3) leads to the following:

$$\delta = -D\frac{V_{\text{in}}}{V_{\text{o}}}. \tag{3.21}$$

At the zero-state, the inductor current maintains zero and there is no current flowing to the load which seems that the converter falls into sleeping interval.

As analyzed above, no current flows to the diode and load during the on-state. The output current (I_{o}) is equal to the average of the diode current. During the off-state, the diode current is equal to the inductor current. Therefore, I_{o} can be written as following:

$$I_{\text{o}} = \delta\frac{I_{\text{LP}}}{2}. \tag{3.22}$$

Based on Eqs. (3.19), (3.21) and (3.22), V_{o} is obtained as following:

$$V_{\text{o}} = -\frac{D^2T}{2LI_{\text{o}}}V_{\text{in}}^2. \tag{3.23}$$

Compared to Eq. (3.15), the relationships of the input/output voltage for DCM are much more complicated than that in CCM.

Additionally, considering $P_{\text{o}} = |V_{\text{o}}I_{\text{o}}|$, Eq. (3.23) can be rewritten as following:

$$P_{\text{o}} = \frac{D^2T}{2L}V_{\text{in}}^2. \tag{3.24}$$

Assuming it is an ideal DC-DC converter, then $P_{\text{in}} = P_{\text{o}}$ is formed. Combining Eqs. (3.9) and (3.24), it is obvious that:

$$R_{\text{in}} = \frac{2L}{D^2T}. \tag{3.25}$$

If f_{sw} is the frequency of the V_{gate}, then, $f_{\text{sw}} = \frac{1}{T}$. Hence, Eq. (3.25) can be rewritten as following:

$$R_{\text{in}} = \frac{2Lf_{\text{sw}}}{D^2}. \tag{3.26}$$

As seen from Eq. (3.26), the input resistance R_{in} is independent of the input voltage V_{in} and the load resistance R_{L} with respect to the buck-boost converter in DCM. The input resistance R_{in} is only decided by the inductance L, switching frequency f_{sw} and the duty-on ratio D.

B). Buck converter

For the same reason, the variation of the inductor current should be zero among the on-state and off-state. Then δ can be given by following:

$$\delta = D\frac{V_{\text{in}} - V_{\text{o}}}{V_{\text{o}}}. \tag{3.27}$$

The current always flows through the inductor and load both at on-state and off-state as shown in Fig. 3.2(a). Therefore, the output current I_{o} should be equal to the average of the inductor current I_{L}.

$$I_{\text{o}} = \frac{1}{2}(D + \delta)I_{\text{LP}}. \tag{3.28}$$

Applying Eqs. (3.5) and (3.27) to Eq. (3.28) leads to the following:

$$V_{\text{o}} = \frac{1}{\frac{2LI_{\text{o}}}{D^2T} + V_{\text{in}}}V_{\text{in}}^2. \tag{3.29}$$

Similarly, Eq. (3.29) can be rewritten as

$$R_{\text{in}} = \frac{2Lf_{\text{sw}}}{D^2}\frac{1}{1 - \frac{V_{\text{o}}}{V_{\text{in}}}}. \tag{3.30}$$

Therefore, the input resistance R_{in} is complicated and depended on the proportion of output and input voltages.

C). Boost converter

Similarly, as for boost converter, the V_{o} and R_{in} also can be expressed as following:

$$V_{\text{o}} = (\frac{D^2T}{2LI_{\text{o}}}V_{\text{in}} + 1)V_{\text{in}}, \tag{3.31}$$

$$R_{\text{in}} = \frac{2Lf_{\text{sw}}}{D^2}(1 - \frac{V_{\text{in}}}{V_{\text{o}}}). \tag{3.32}$$

Finally, we summarize the relationships of the input/output resistances and input/output voltages for these three basic topologies DC-DC converters in DCM as shown in Table 3.2. Referring to the Table 3.1, R_{in} of three topologies DC-DC converters in CCM are all relative to the load resistance R_{L}. In the DCM case, R_{in} of buck and boost converters are both relative to the proportion of V_{in} and V_{o}. However, the R_{in} of the buck-boost converter is independent of V_{in}, V_{o} and R_{L}. If the inductance L, switching frequency f_{sw}, duty-on ratio D are decided, the R_{in} of the buck-boost converter can be calculated and it will be constant. Therefore,

Table 3.2: Input/output resistance and voltage relationships of three basic topologies DC-DC converters in DCM.

Topology	Voltage	Resistance	Variation of R_{in}
buck	$V_{\text{o}} = \frac{1}{\frac{2LI_{\text{o}}}{D^2T}+V_{\text{in}}}V_{\text{in}}^2$	$R_{\text{in}} = \frac{2Lf_{\text{sw}}}{D^2}\frac{1}{1-\frac{V_{\text{o}}}{V_{\text{in}}}}$	depends on $\frac{V_{\text{o}}}{V_{\text{in}}}$
boost	$V_{\text{o}} = (\frac{D^2T}{2LI_{\text{o}}}V_{\text{in}}+1)V_{\text{in}}$	$R_{\text{in}} = \frac{2Lf_{\text{sw}}}{D^2}(1-\frac{V_{\text{in}}}{V_{\text{o}}})$	depends on $\frac{V_{\text{o}}}{V_{\text{in}}}$
buck-boost	$V_{\text{o}} = -\frac{D^2T}{2LI_{\text{o}}}V_{\text{in}}^2$	$R_{\text{in}} = \frac{2Lf_{\text{sw}}}{D^2}$	independent of V_{in} and R_{L}

a buck-boost converter operated in DCM is chosen for impedance matching in the rectenna circuit.

3.3 Externally-powered DCM Buck-boost Converter

3.3.1 Input Resistance R_{in}

The buck-boost converter in DCM exhibits constant input resistance despite the load resistance as mentioned above. However, it inverts the polarity of input and output voltages. Therefore, we design an inverting DCM buck-boost converter which inputs a negative voltage to get a positive output voltage. The proposed inverting DCM buck-boost converter is shown in Fig. 3.4(a) where the control-pulse circuit consists of a 5 V power source, a low-frequency (LH) oscillator, a high-frequency (LH) oscillator and a MOSFET gate driver. The LF and HF oscillators respectively generate the LF, HF pulse waves which are combined into the gate control pulse V_{gate}. Through the gate driver, the V_{gate} controls the N-MOS's gate on or off. Here, the input port of the converter is expected to be connected to the output of the rectenna (or rectifier). Figure 3.4(b) shows the operating waveform of the proposed converter where I_{in}, I_{L}, I_{D} are input current, inductor current, diode current, respectively, and V_{gate}, V_{h}, V_{ι} are gate control pulse wave, HF pulse wave, LF pulse wave, respectively. An additional 5 V power source powers the LF oscillator and gate driver, and the HF oscillator is powered

by the LF oscillator. It means that the V_{gate} is as same as the V_{h} when the V_{ι} is on while the V_{gate} is off when the V_{ι} is off. As seen from the waveform, there is no current flowing though the circuit at the off-state of the LF oscillator. It seems that the LF oscillator compels the circuit to be sleep state at an interval then the current becomes discontinuous.

The operating condition can be classified into two states such as a work-state and a sleep-state. The work-state is when the LH pulse is on. At work-state, there are three operating states as same as mentioned in Chapter 3.2.2. The sleep-state is when the LH pulse is off. At sleep-state, there is no current flowing through the circuit. As shown in Fig. 3.4(b), the input current is equal to the inductor current only at the on-state. If I_{inave} is the average input current, then:

$$I_{\text{inave}} = \frac{I_{\text{LP}} D_{\text{h}} D_{\iota}}{2}, \tag{3.33}$$

where D_{h}, D_{ι} are the high, low frequency pulse duty-on ratio. Referring to Eq. (3.19), the peak inductor current I_{LP} can be written as following:

$$I_{\text{LP}} = \frac{V_{\text{in}} D_{\text{h}} T_{\text{h}}}{L}, \tag{3.34}$$

where the T_{h} is the period of the HF pulse. Furthermore, input power P_{in} satisfies the following equation:

$$P_{\text{in}} = V_{\text{in}} I_{\text{inave}}. \tag{3.35}$$

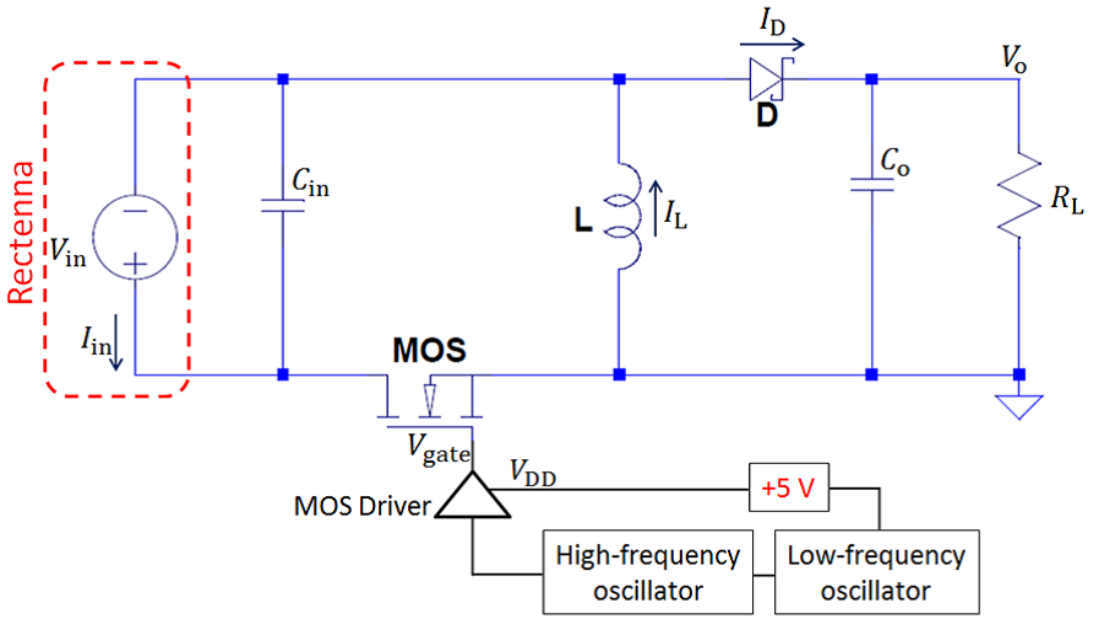
Combining Eqs. (3.9), (3.33), (3.34) and (3.35), the input resistance R_{in} is given as follows:

$$R_{\text{in}} = \frac{2L}{D_{\text{h}}^2 D_{\iota} T_{\text{h}}}. \tag{3.36}$$

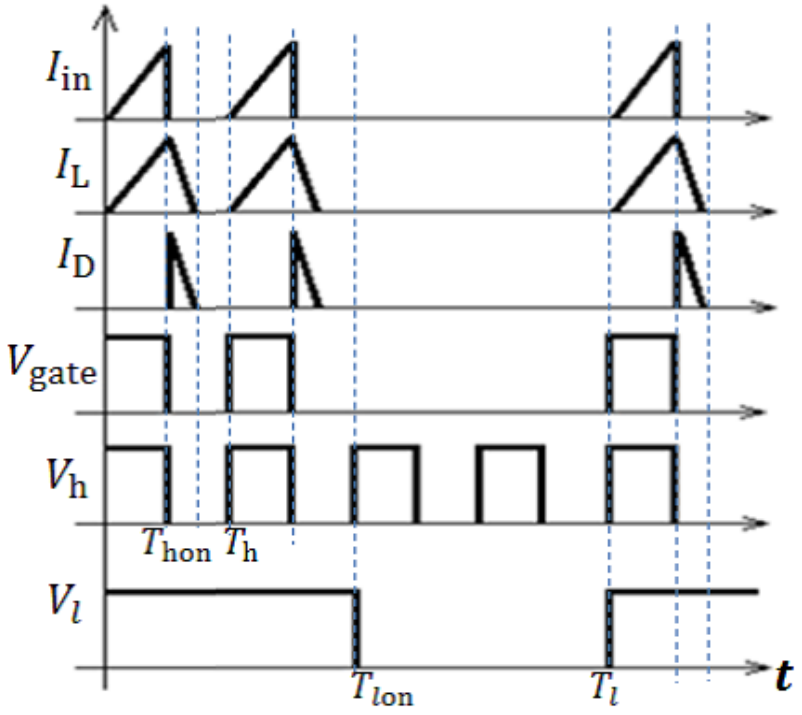
Let f_{h} represent the frequency of the HF pulse, applying $T_{\text{h}} = \frac{1}{f_{\text{h}}}$ to Eq. (3.36) leads to the following:

$$R_{\text{in}} = \frac{2L f_{\text{h}}}{D_{\text{h}}^2 D_{\iota}}. \tag{3.37}$$

Hence, the R_{in} of the proposed inverting DCM buck-boost converter is decided by the inductance L , low, high frequency pulse duty-on ratio D_{ι}, D_{h} and the high frequency f_{h}.



(a)



(b)

Figure 3.4: Inverting DCM buck-boost converter. An extra +5 V power source is required for the control-pulse circuit. Input a negative voltage to obtain a positive output voltage. (a) Structure. (b) Operating waveforms.

Table 3.3: Components of the proposed inverting DCM buck-boost converter.

Component	Maker	Part NO.	Description
High-frequency oscillator	Linear Technology	LTC6900	$I_{SS} = 0.5$ mA ($f = 3$ MHz); $V_{SSmin} = 2.7$ V
Low-frequency oscillator	Linear Technology	LTC6992	I_{SS}= 115 μA ($f = 100$ Hz); $V_{SSmin} = 2.25$ V
N-MOS	Fairchild	FDS6298	$V_{DSS} = 30$ V; $R_{DS(on)} = 12$ mΩ ($V_{GS} = 4.5$ V)
Schottky Diode	Toshiba	CRS01	$V_{RRM} = 30$ V; V_{FM}= 0.36 V
MOSFET Driver	Microchip	TC1427	$V_{DDmin} = 4.5$ V
Inductor	Panasonic	ELC18B102L	$L = 1$ mH; $R_{dc} = 0.39$ Ω; I_{dc} = 1.1 A
Capacitor	Rubycon	Electrolytic Capacitors	$C_{in} = 47$ μF ($V_{rated} = 10$ V); $C_o = 100$ μF ($V_{rated} = 50$ V)

3.3.2 Simulation and Experimental Results

In this buck-boost converter, set $L = 1$ mH, $D_h = 0.5$, $D_\iota = 0.6$, and $f_h = 10$ kHz, so that the input resistance of the converter R_{in} becomes 133 Ω, as calculated using Eq. (3.37). Table 3.3 lists the components used in the buck-boost converter, and Fig. 3.5 shows a photograph of the buck-boost converter, which uses dielectric substrate as a printed circuit board (PCB).

In order to verify the theory as mentioned above, we simulate the converter by LTspice and conduct an experiment on it. The structure of the experiment circuit is shown in Fig. 3.6, where a negative voltage supplied by a DC power source is the input voltage. A +5 V power source supplies the power for the control-pulse circuit. An ammeter and a voltmeter measure the input current and the output voltage, respectively. We measure the efficiency and the input resistance of the converter with an input voltage from 0.8 to 5 V, where the input power is from 5 to 187 mW. Figure 3.7 shows the simulation and experimental results for input

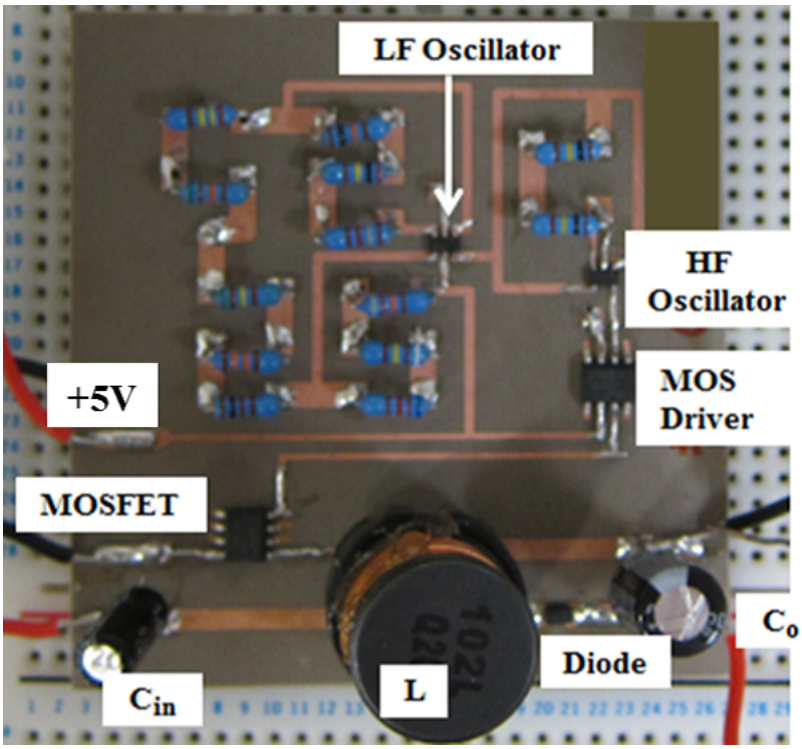


Figure 3.5: Photograph of the inverting DCM buck-boost converter.

resistance R_{in} versus the load resistance R_L from 100 to 5000 Ω for the input voltage of 1.2 and 3 V. According to Fig. 3.7, the simulation results almost agree well with the theory of 133 Ω. The experimental R_{in} is near the calculated value of 133 Ω, but it is slightly lower for a low load resistance when the R_L is smaller than 400 Ω. This is because the inductor current becomes continuous at the beginning of the switching-on period of the MOSFET at a low load resistance. However, the equation of the input resistance (Eq. (3.37)) is calculated as a discontinuous conduction mode. Thus, the measured input current is slightly larger than the calculated value. Consequently, the input resistance is smaller than the calculated value. Figure 3.8 shows the R_{in} versus the input power from 2 to 187 mW for the load resistance of 1000 Ω. According to Fig. 3.8, the experimental R_{in} almost agrees with the simulation results, and it is varying in the range of 123 to 133 Ω as the input power is changing from 5 to 187 mW. Therefore, according to Figs. 3.7 and 3.8, as expected, the designed buck-boost converter has constant input resistance despite the load resistance and the input power.

Furthermore, the efficiency of the converter is also measured. Here, the efficiency does not include the power dissipated by the control circuit. Similarly, Figure 3.9 shows the measured efficiency of the converter for load resistances from 100 to 5000 Ω for the input voltage of 1.2 and 3 V, where the input power are 12

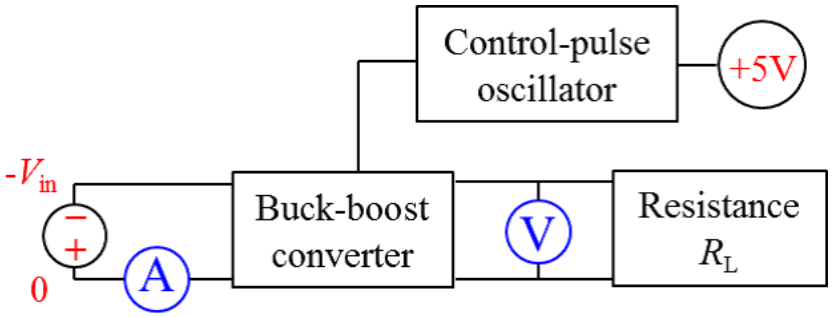


Figure 3.6: Measurement setup of the externally-powered buck-boost converter. An extra +5 V power source is required.

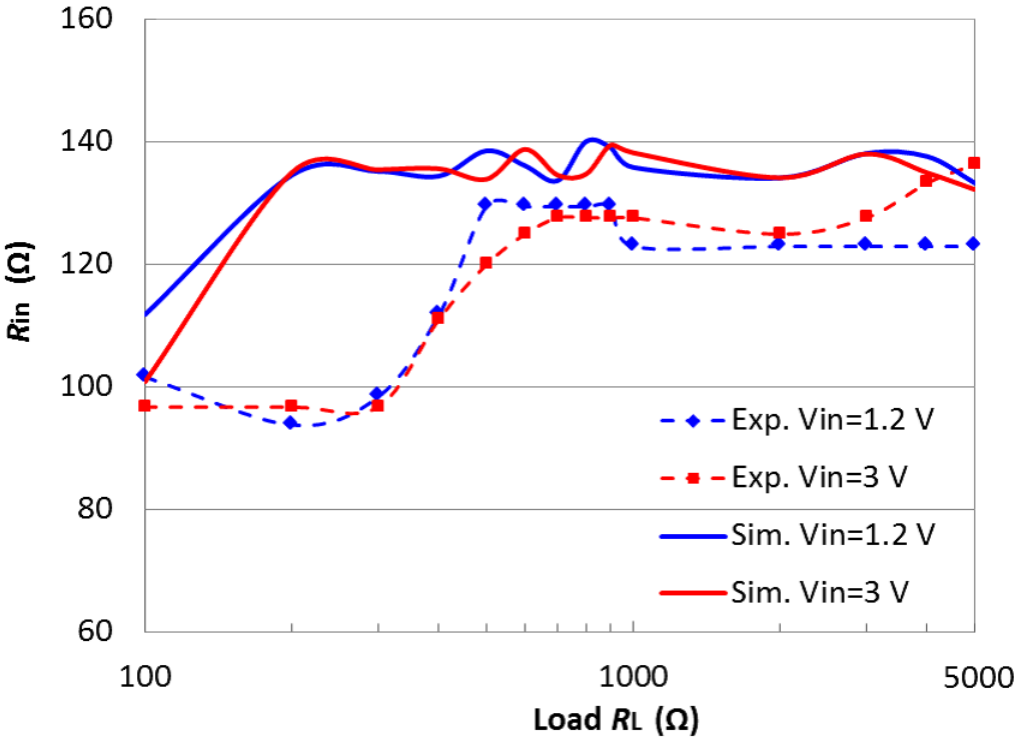


Figure 3.7: Simulation and experimental results of the input resistance R_{in} vs. load resistance for an input voltage of 1.2 or 3 V.

and 70 mW, respectively. The efficiency is almost greater than 80 % for a wide range of loads from 200 to 5000 Ω at the input voltage of 3 V. The efficiency is a little low at load resistance of 100 Ω because the load resistance is smaller than the input resistance and the input voltage is reduced by the converter. At this moment, the converter works as a buck converter, where the output voltage is lower than the input voltage. The efficiency decreases under a large load resistance because the output current is shunted by the output capacitor. The current flowing through the output capacitor will increase as the load resistance is increased. Figure 3.10 shows the measured efficiency of the converter for input power from

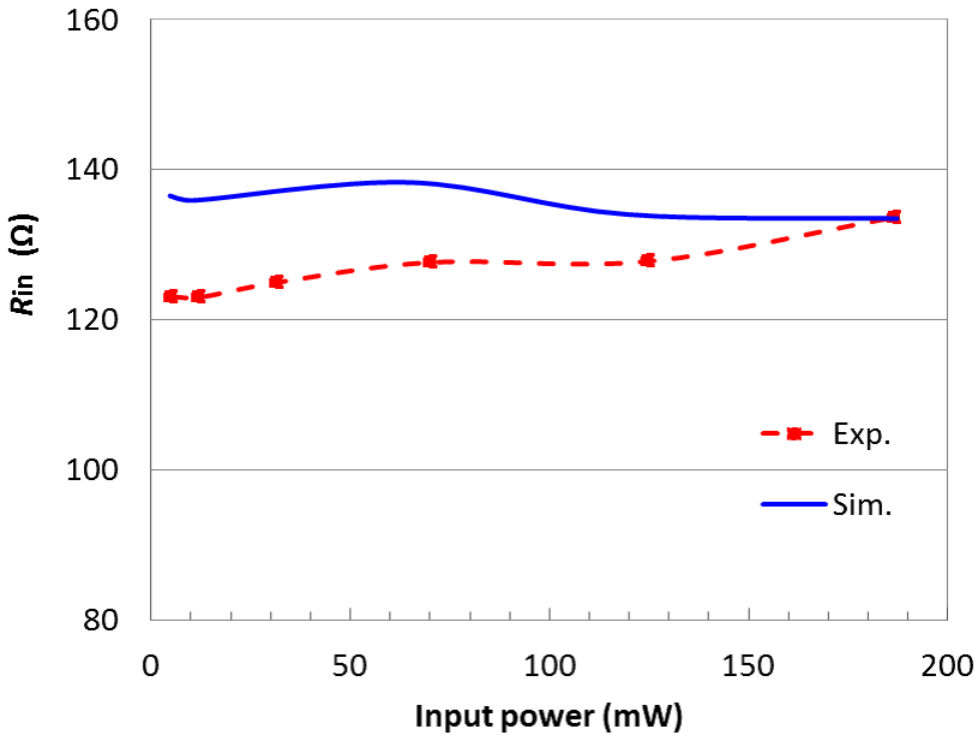


Figure 3.8: Simulation and experimental results of the input resistance R_{in} vs. input power for a fixed load resistance of 1000 Ω.

5 to 187 mW for a fixed load resistance of 1000 Ω. The efficiency is increasing lightly as the input power is increasing. In this case, the measured efficiency is greater than 85 % for this wide range of input powers. Additionally, since ideal capacitors and wires are used in the simulation, the measured efficiency is almost 5 % lower than the simulation results which both can be seen from Figs. 3.9 and 3.10.

3.4 Rectifying Circuit with Negative Output Voltage

As mentioned in Chapter 2.3, there are two types of dc pass filters such as a $\lambda/4$ line plus a capacitor [58] and a $\lambda/4$ line plus some open stubs in single shunt rectifier. The authors obtained a higher experimental efficiency using open stubs instead of capacitor [59]. We also utilize a $\lambda/4$ line plus four open stubs as a dc pass filter as shown in Fig. 3.11(a). The impedance of the dc pass filter Z_{dc}, opposite to the class-F^{-1} mode [68], presents zero for even harmonics and

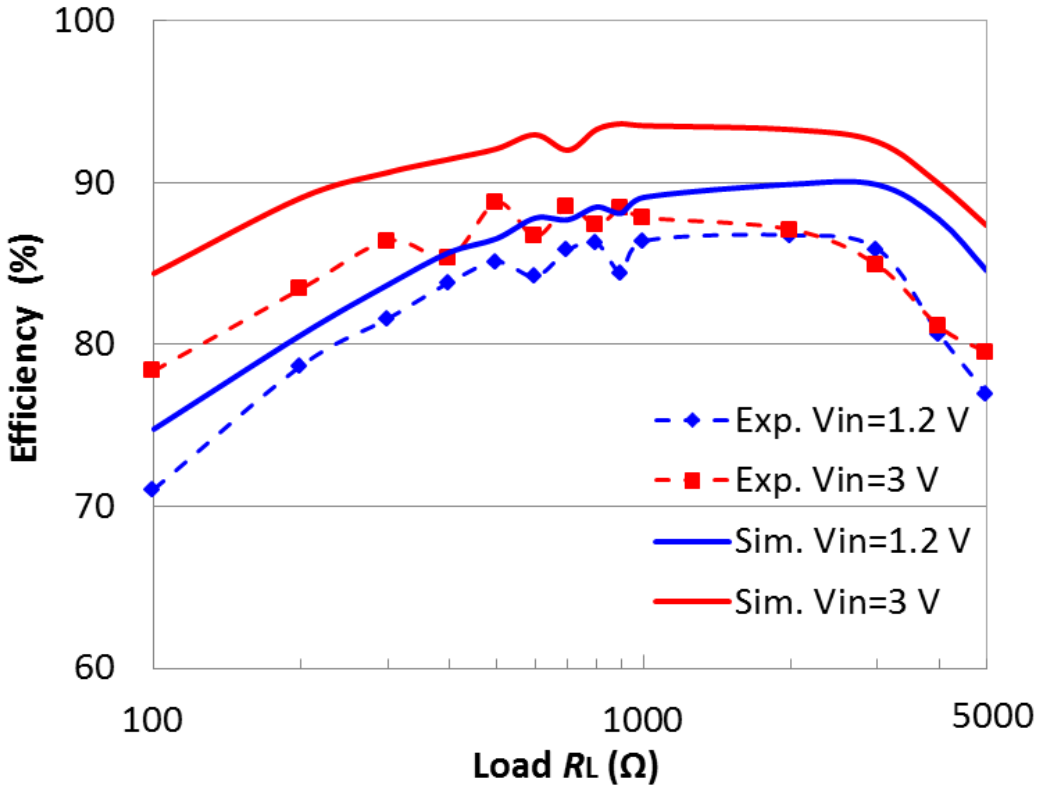


Figure 3.9: Simulation and experimental results of the efficiency vs. load resistance for an input voltage of 1.2 or 3 V.

infinity for odd harmonics. In other words, all even harmonics are terminated in short circuits and all odd harmonics are terminated in open circuits. The set of harmonic terminations is the same as for a class-F amplifier [69], thus this rectifier is referred to as a class-F rectifier. In fact, the impedance of the dc pass filter, consisted of a $\lambda/4$ line and a capacitor in shunt, also presents a class-F characteristic which is analyzed in [59]. Due to the parasitic components of a real capacitor, the combination of a $\lambda/4$ line and a capacitor can not satisfy the demand of the dc pass filter characteristic for high harmonica frequencies.

The designed buck-boost converter inverts the polarity of the input voltage and the output voltage, which is described in Chapter 3.3. Hence, unlike other rectifiers, we reverse the polarity of the diode in order to obtain a negative output voltage from the rectifier. There is a little discussion and comparison about the diode topologies in single, two-diode package and two anti-parallel diodes in [57] which indicates that two-diode rectifier would result in higher efficiency for high power levels. Additionally, the rectifier using two-diode parallel HSMS2865 or three-diode parallel HSMS286L obtained a higher efficiency in simulation results

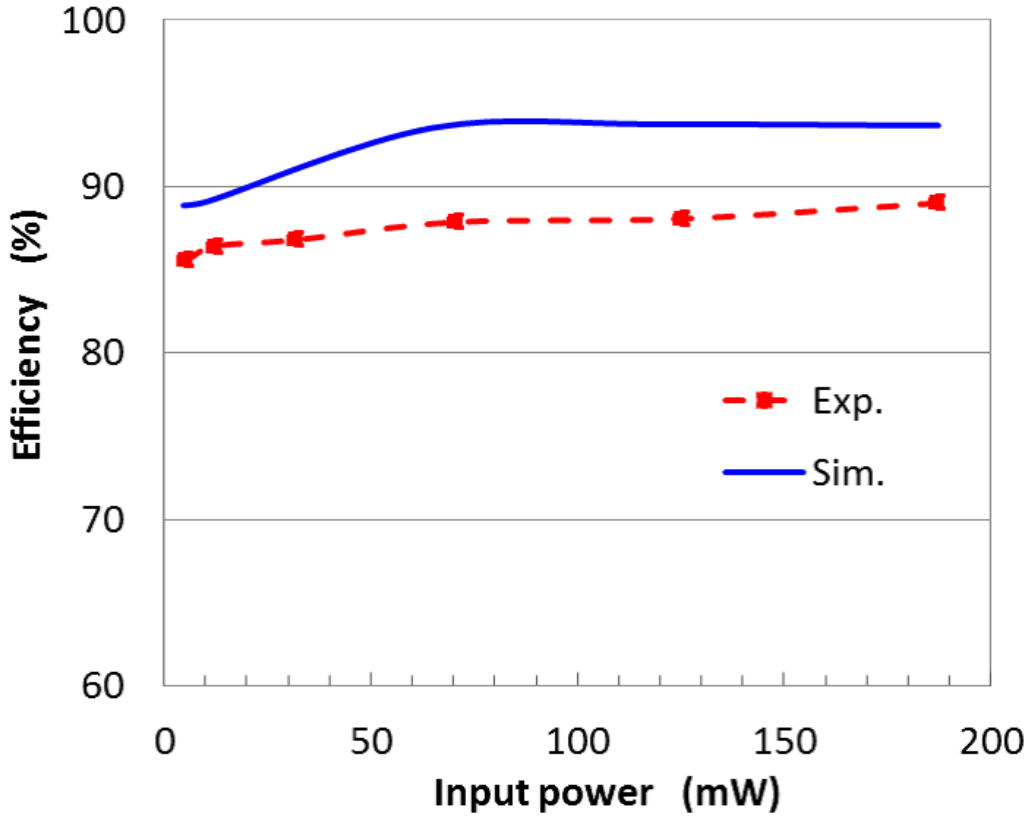
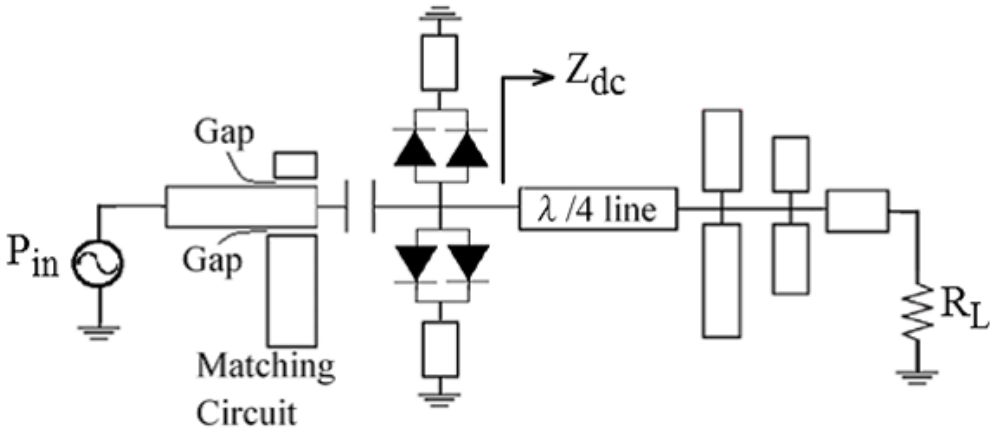


Figure 3.10: Simulation and experimental results of the efficiency vs. input power for a fixed load resistance of 1000 Ω.

in Chapter 2.3. However, as seen from Fig. 2.2, the physical size of HSMS2865 (4 pins) and HSMS286L (6 pins) are too wide to limit the characteristic impedance of the connected microstrip line. Therefore, considering the physical size and the parallel diodes, we choose two HSMS286F as a rectifying diode. HSMS286F diode, consisting of two HSMS2860 diodes mounted into a single surface mount package, has 3 pins because of a common cathode pin. Table 3.4 lists the components used in the rectifier and Fig. 3.11(b) shows the photograph of the designed rectifier.

We first simulate the rectifying circuit using ADS2011. We optimize the whole rectifying circuit to gain the highest efficiency not just considering the impedance matching with the diodes but also reducing the dissipation power of diodes in the simulation. Then, we fabricate it and test it with 2.45 GHz microwave power. The simulation and experimental results of the rectifier circuit are shown in Fig. 3.12.

Figure 3.12(a) shows the RF-DC conversion efficiency versus load resistances ranging from 50 to 1000 Ω, which represents a typical efficiency characteristic of



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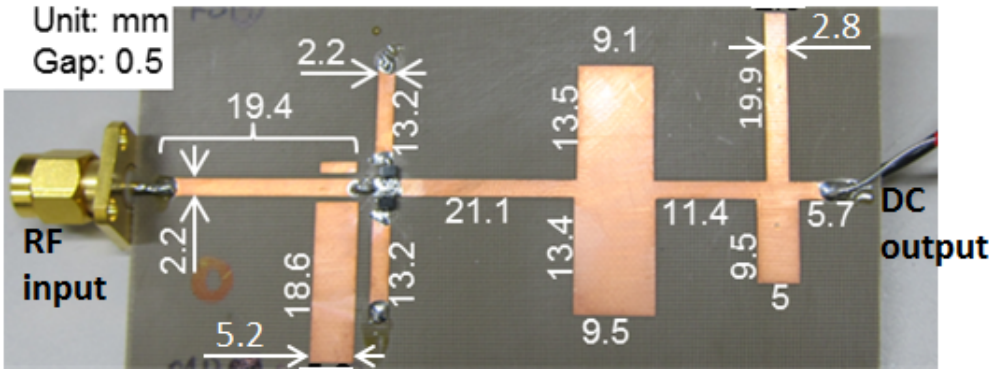
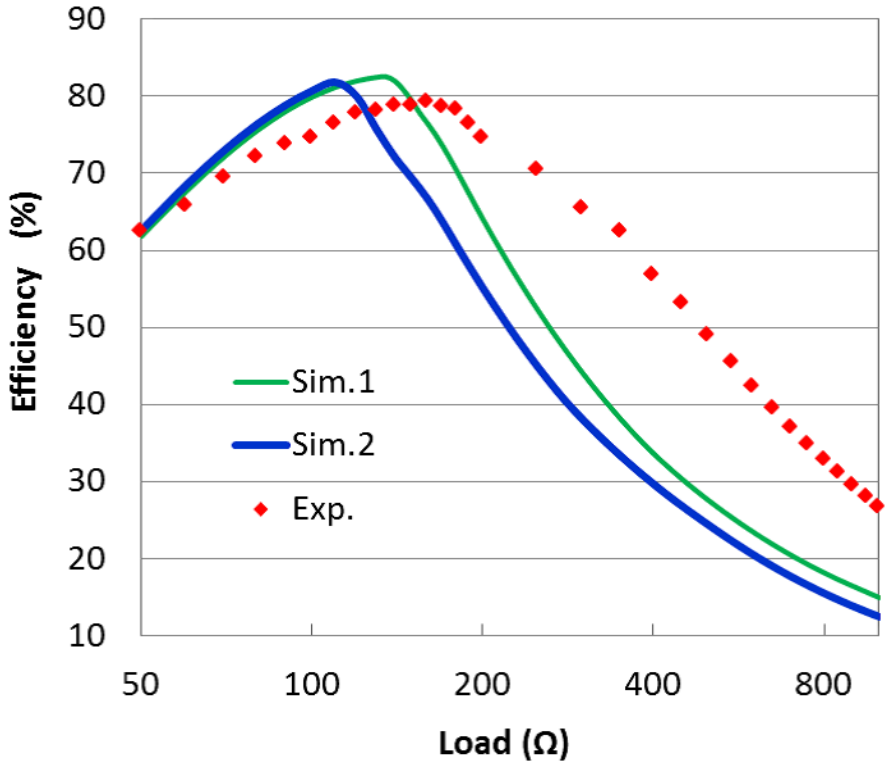


(b)

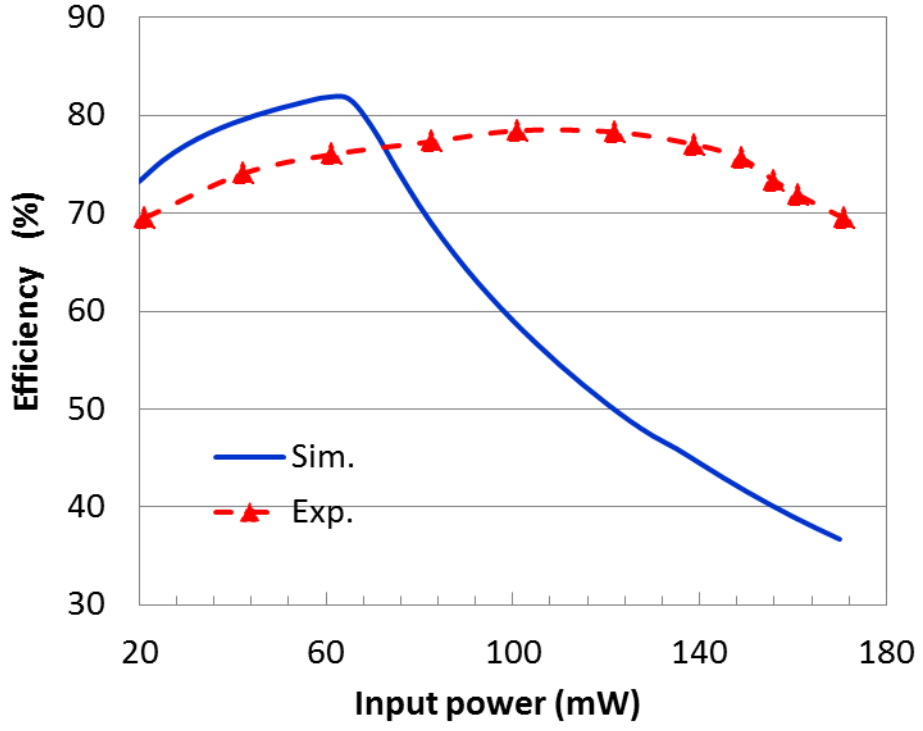
Figure 3.11: Proposed negative output voltage rectifier. (a) Structure. (b) Photograph.

the rectifying circuit. The two solid curves are simulation results and the dot curve is the experimental results. Sim.1 gets the maximum simulation efficiency of 82.5 % at the input power of 69 mW. However, the maximum experimental efficiency is 79.2 % at input power of 82 mW as shown in Exp. curve. Then, when we set the input power as 82 mW in simulation, the maximum efficiency of 81.8 % is obtained (Sim.2 curve). The detailed results of these three curves are summarized in Table 3.5. We obtain the maximum efficiency of 79.2 % at an optimal load of 160 Ω. In particular, the measured efficiency difference between an optimal load of 160 and 1000 Ω is approximately 53 %, which clearly indicates that the efficiency of the rectifier depends on the load resistance.

Figure 3.12(b) shows the simulation and experimental efficiency versus the



(a)



(b)

Figure 3.12: Simulation and experimental results of the proposed negative output voltage rectifier. (a) Efficiency vs. load resistance for a constant input power of 82 mW. (b) Efficiency vs. input power for a constant load of 150 Ω.

Table 3.4: Components of the negative output voltage rectifier.

Component	Maker	Part NO.	Description
DC block capacitor	Murata	GRM1882C1-H331JA01D	$C = 330$ pF; $V_{\text{rated}} = 50$ V
HF diode	Avago	HSMS286F	Refer to Table 2.2
Dielectric substrate	Nippon pillar	NPC-F260	$\varepsilon_{\text{r}} = 2.53$; $\tan\delta = 0.0018$; $t = 18$ μm; H = 0.8 mm

Table 3.5: Comparison of simulation and experimental results of the rectifier.

	Input power	Efficiency	Reflection	Optimal load
Sim.1	69 mW	82.5 %	1.6 %	135 Ω
Exp.	82 mW	79.2 %	4.9 %	160 Ω
Sim.2	82 mW	81.8 %	3.4 %	110 Ω

input power ranging from 20 to 170 mW at the load of 150 Ω. The efficiency is changing with the input power while the maximum difference in efficiency is only 10 % for this range of input powers. In simulation, the maximum efficiency is obtained at the input power of 60 mW, while the efficiency drops rapidly with the input power exceeding 60 mW because the reverse voltage is over the breakdown voltage of the diode. However, in experiment, the efficiency gets much higher even though the input power increases far over 60 mW. A similar phenomena can be seen from Fig. 3.12(a), the experimental efficiency is higher than the simulation one when the load resistance exceeds the optimal load. This is because the real breakdown voltage is higher than the value (7 V) given by the SPICE parameters. A high precision HF diode model is required in ADS simulation.

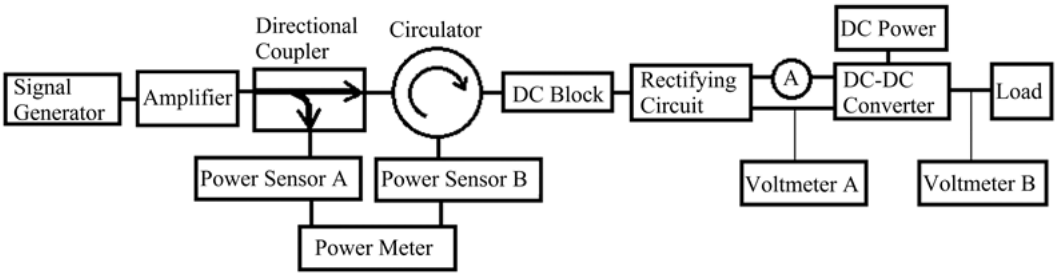
Table 3.6: Efficiency and load relationships of the rectifier and buck-boost converter in experimental results.

	R_L	R_{in}	Efficiency
RF-DC rectifier	130 Ω		78 %
buck-boost converter	400-5000 Ω	130 Ω	80 %
RF-DC-DC circuit	400-5000 Ω		predicted value 62 %

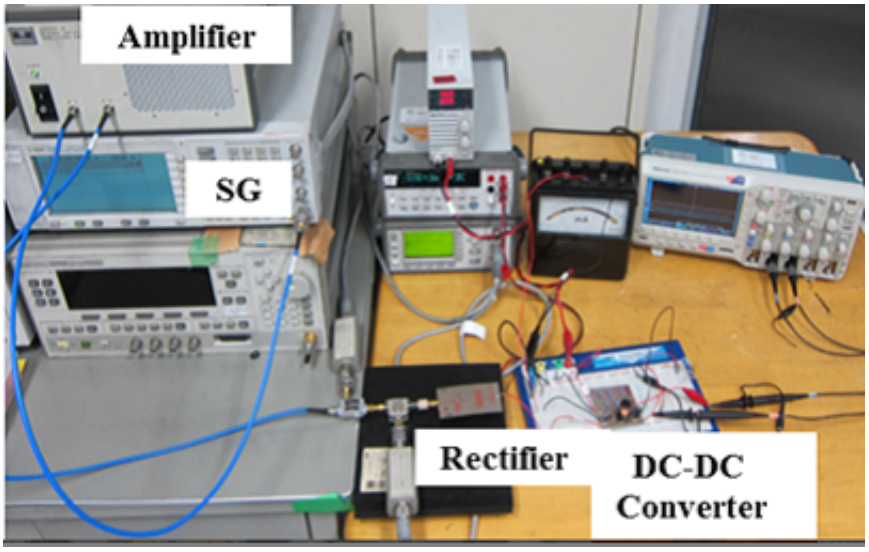
3.5 Experiment on Externally-powered RF-DC-DC Circuit

The experiment on the RF-DC-DC circuit consisted of the rectifier and the buck-boost converter designed above is conducted. Figure 3.13(a) is the experimental architecture for measuring the efficiency of the RF-DC-DC circuit and Fig. 3.13(b) is the experiment photograph. The microwave power is first regulated by the RF-DC rectifier, and the DC output voltage of the RF-DC rectifier is again regulated by the buck-boost converter. The input resistance of the DC-DC converter seen from the left side is also the output impedance of the rectifier. Summarize the efficiency and load relation of the rectifier and the buck-boost converter designed above in Table 3.6. The designed R_{in} of the converter is 133 Ω which is slightly different with the optimal load (160 Ω) of the rectifier, while the efficiency of the rectifier at 133 Ω is only 1.2 percentage points lower than that at the optimal load. Thus, as expected, the buck-boost converter can track the maximum efficiency of the rectifier and the overall efficiency of the RF-DC-DC circuit is expected to be steady and over 62 %.

Following Fig. 3.13(a), we use a 2.45 GHz microwave power generated from a signal generator. Power sensor A measures the input power of the rectifier. Through the circulator, power sensor B measures the reflection power at the input point of the DC block. Here, we use a DC block to prevent the DC voltage from passing through the rectifier, which also can prevent the positive output pole of the rectifier from being grounded, because the positive input pole of the buck-boost converter is not allowed to become grounded. Referring to Fig. 3.4, if the



(a)



(b)

Figure 3.13: Experimental setup for measuring the efficiency of the RF-DC-DC circuit. (a) Block diagram. (b) Experiment photograph.

positive input pole of the converter is grounded, the MOSFET will be shorted and the converter can not act as a buck-boost converter. An ammeter is used to measure the input current of the converter, and voltmeters A and B are used to measure the output voltage of the rectifier and the output voltage of the converter, respectively. We use DC power for the control circuit to supply voltage to the LF oscillator and the gate driver in the experiment. Before reporting the experimental results, the tracking efficiency η_{track} is defined as follows:

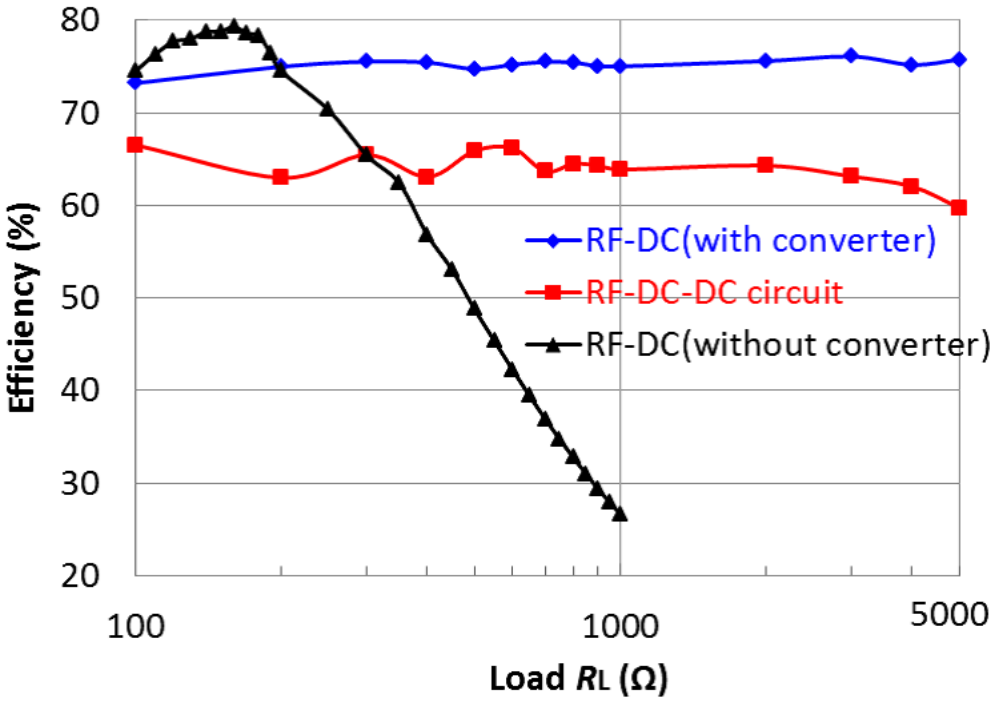
$$\eta_{\text{track}} = \frac{\eta'_{\text{rf-dc}}}{\eta_{\text{max(rf-dc)}}} \times 100\,\%, \tag{3.38}$$

where $\eta'_{\mathrm{rf-dc}}$ is the conversion efficiency of the RF-DC rectifying circuit when it is connected to the buck-boost converter, and $\eta_{\mathrm{max(rf-dc)}}$ is the maximum conversion efficiency of the RF-DC rectifying circuit at the optimal load without a buck-boost converter.

Figure 3.14 shows the measured efficiency versus the load resistance from 100 to 5000 Ω and an input power from 20 to 100 mW. Figure 3.14(a) compares the efficiency of the RF-DC-DC circuit with the rectifying circuit described in Chapter 3.4 under the input power of 82 mW. The blue curve, RF-DC (with converter), is the efficiency of only the rectifying circuit when it is connected to the buck-boost converter. The black curve, RF-DC (without converter), is the efficiency of the rectifying circuit alone when it is not connected to the buck-boost converter. The black curve decreases dramatically when the load exceeds the optimal load and the efficiency is only 26 % at the load of 1000 Ω. In comparison, the conversion efficiency of the rectifying circuit connected with the converter is almost steady at 75 %, which is near the maximum power point of the rectifying circuit, even though the load resistance is changed from 100 to 5000 Ω. Furthermore, as expected, the overall efficiency of the RF-DC-DC circuit is over 60 % in a wide range of load resistances from 100 to 5000 Ω, as indicated by the red curve. Compared to the rectifying circuit without a converter, the overall efficiency is lower than that when the load is below 300 Ω. However, the overall efficiency is much higher than that when the load exceeds 300 Ω. Under the large load resistance, the overall efficiency decreases slightly because the efficiency of the buck-boost converter decreases as the load increases.

Figure 3.14(b) shows the overall efficiency versus the input power changing from 20 to 100 mW at the load resistance of 1000 Ω. Note that the variation of the overall efficiency is small and the overall efficiency is over 55 % in this power range.

Additionally, we also check the tracking efficiency and the input resistance R_{in} of the buck-boost converter when it is utilized in the RF-DC-DC circuit as shown in Fig. 3.15. Figure 3.15(a) shows that the tracking efficiency η_{track} remains over 92 % over the entire range of loads from 100 to 5000 Ω and the input resistance R_{in} is approximately near to the designed value of 133 Ω. The coherent variation trend of two curves can be seen from Fig. 3.15(a). This is because the $\eta'_{\mathrm{rf-dc}}$ is increasing when the R_{in} is close to 133 Ω, which leads to η_{track} increasing according to Eq. (3.38). Similarly, Fig. 3.15(b) presents η_{track} and R_{in} with input power ranging



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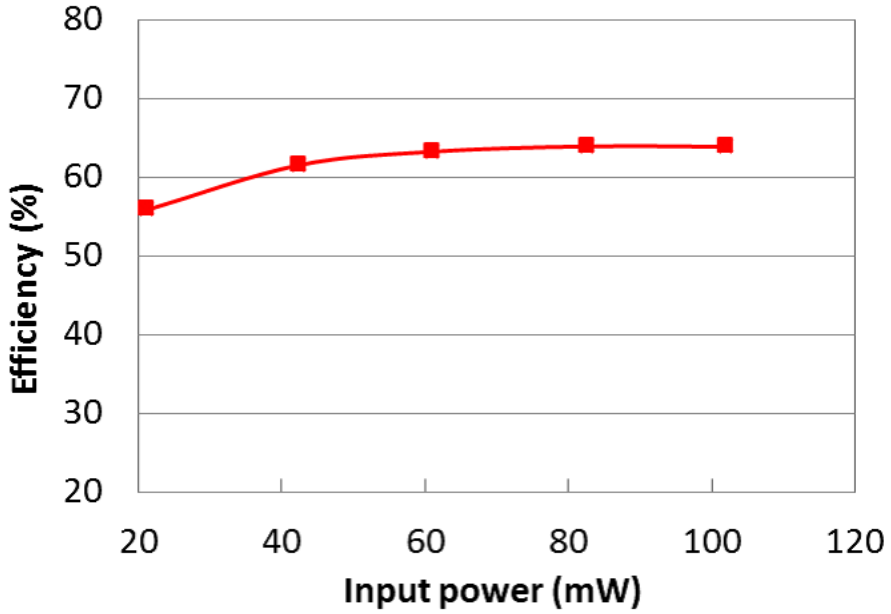


(b)

Figure 3.14: Measured overall efficiency of the RF-DC-DC circuit. (a) Efficiency comparison of the RF-DC-DC circuit and the general rectifier for a constant input power of 82 mW. (b) Overall efficiency vs. input power for a constant load of 1000 Ω.

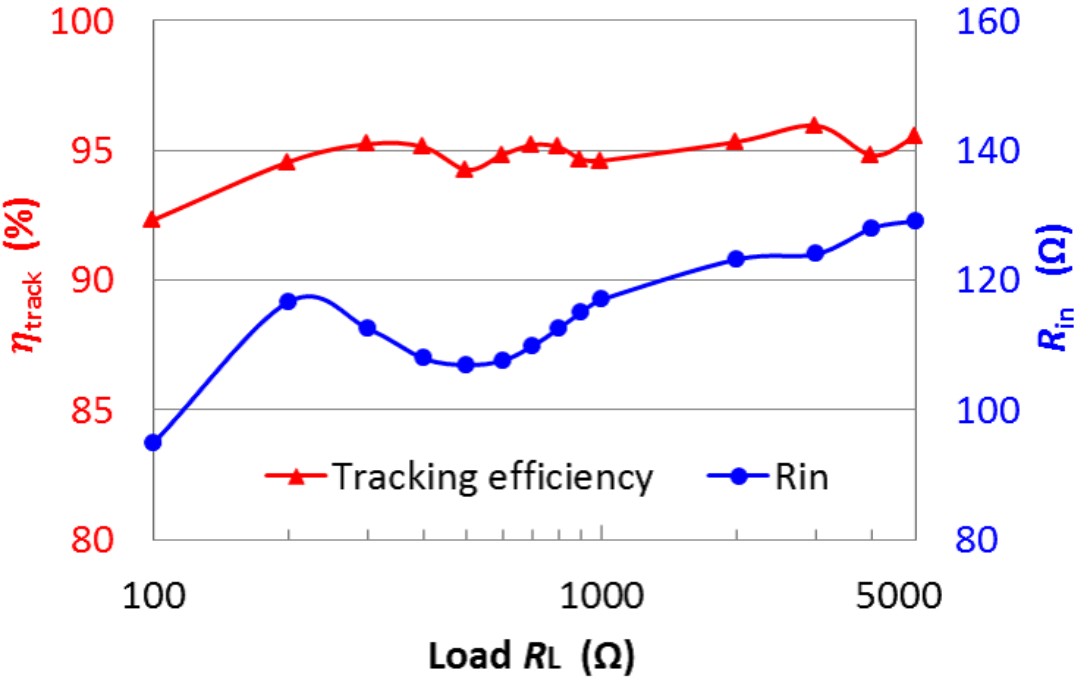
from 20 to 100 mW at the load resistance of 1000 Ω. Note that the tracking efficiency is slightly increasing as the input power increases. The reason is as follows. The input resistance of the converter is close to the optimal load of the rectifier as the input power is increased, which leads to increase in the efficiency of the rectifier ($\eta'_{\mathrm{rf-dc}}$). However, the maximum difference of the tracking efficiency is about 2 percentage points in this power range.

Figure 3.16 compares the output voltage versus load resistance at the input power of 82 mW. The red curve is the output voltage of the RF-DC-DC circuit, which is increasing as the load resistance increases. The black curve is the output voltage of the rectifier when it is not connected to the converter. Compared to it, the blue curve is the output voltage of the rectifier when it is connected to the converter, where the voltage is almost steady at 2.7 V. Note that the RF-DC-DC circuit obtains a higher output voltage than the usual rectifier at a same load resistance. Here, the relation of three voltage curves agree well with the simulation results reported in Fig. 2.14.

3.6 Summary

This chapter, firstly, from a new viewpoint, the input/output voltages and input/output resistances relationships of three basic topologies DC-DC converters are discussed. Unlike conventional application, the DC-DC converter is used in resistance conversion instead of voltage conversion. Based on the discussion results, a DCM inverting buck-boost converter, whose input resistance is independent of the load resistance or input power, is chosen for impedance matching in rectenna.

Secondly, we design a DCM inverting buck-boost converter which requires an extra DC power source for powering the control-pulse circuit. The buck-boost converter inputs a negative voltage to obtain a positive output voltage. More importantly, the input resistance of the buck-boost converter match the optimal load of the rectifying circuit to track the maximum efficiency point of the rectifying circuit. As expected, the measured input resistance of the buck-boost converter is approximately 133 Ω, despite the changes of the load resistance and input power. The measured efficiency of the converter exceeds 80 % for a wide range of loads or input powers. Next, we design a rectifying circuit with negative output voltage and peak experimental efficiency of 79.2 %.



(a)

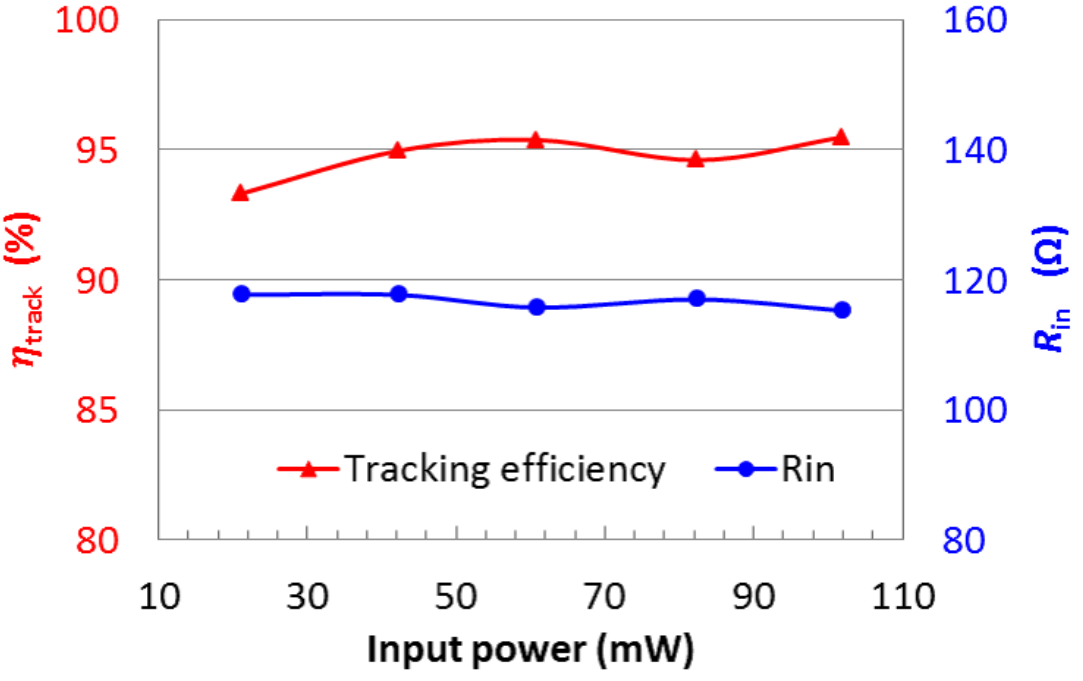


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Figure 3.15: Measured tracking efficiency and input resistance R_{in}. (a) Tracking efficiency and R_{in} vs. load resistance. (b) Tracking efficiency and R_{in} vs. input power.

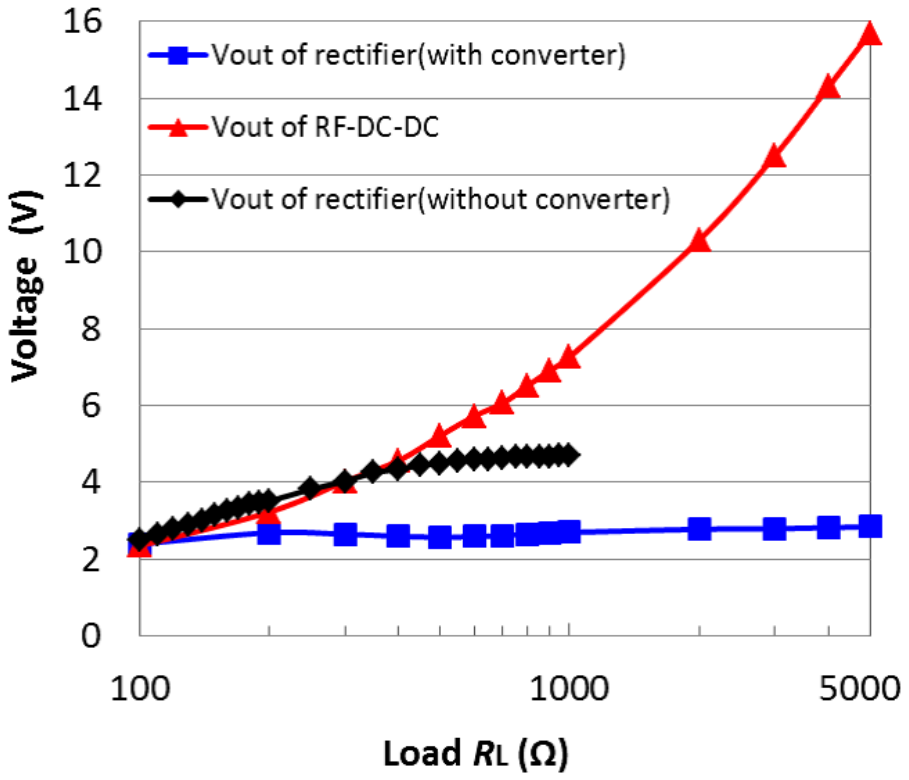


Figure 3.16: Measured output voltage vs. load resistance for a constant input power of 82 mW.

Finally, we verify that the proposed converter successfully tracks the maximum efficiency of the rectifying circuit. The conversion efficiency of the rectifier is approximately steady at 75 %, despite the load resistance varying from 100 to 5000 Ω for an input microwave power of 82 mW. This means that the maximum power tracking efficiency is over 94 %, compared to the maximum efficiency of 79.2 % at the optimal load. Additionally, the overall efficiency of the RF-DC-DC circuit is approximately steady over 60 %, despite the load resistance changing from 100 to 5000 Ω. According to the experimental results, the proposed RF-DC-DC circuit is an effective MPPT method to rectennas.

However, one big disadvantage is that the proposed buck-boost converter requires an additional DC power source for the control-pulse circuit which makes it hard for practical application. Additionally, owing to the loss of the buck-boost converter, the overall efficiency of the RF-DC-DC circuit is much lower than the maximum efficiency of the general rectifying circuit. Therefore, it is necessary to improve the efficiencies both of the rectifying circuit and the converter. Next chapter will give a solution about the problems above.

Chapter 4

Self-powered RF-DC-DC Circuit

4.1 Introduction

Based on the simulation and experiment results of previous discussions, a buck-boost converter in DCM is practicable for impedance matching in rectifying circuit. However, some critical issues such as the lower overall efficiency of the RF-DC-DC circuit and the requirement of an additional DC power source are need to be solved. Therefore, this chapter will solve the issues as mentioned above to improve the RF-DC-DC circuit. To improve the overall efficiency, we design a rectifying circuit with high peak efficiency. Furthermore, the input resistance of the buck-boost converter completely agrees with the optimal load of the rectifying circuit, which increases the tracking efficiency leading to a higher overall efficiency. To remove the additional DC power source, a non-inverting DCM buck-boost converter, whose control-pulse circuit is powered by the output of the rectifying circuit, is proposed. Similarly, an experiment on RF-DC-DC circuit is conducted to verify the advantages of the improved RF-DC-DC circuit.

4.2 High-efficiency Rectifying Circuit

As shown in Fig. 4.1, we also use a class-F rectifier with output filter consisted of a $\lambda/4$ line and four open stubs in this chapter. Compared with the rectifier in Chapter 3.4, the rectifier designed in this chapter has some differences as follows. First, with respect to the input filter, two open stubs are applied for impedance

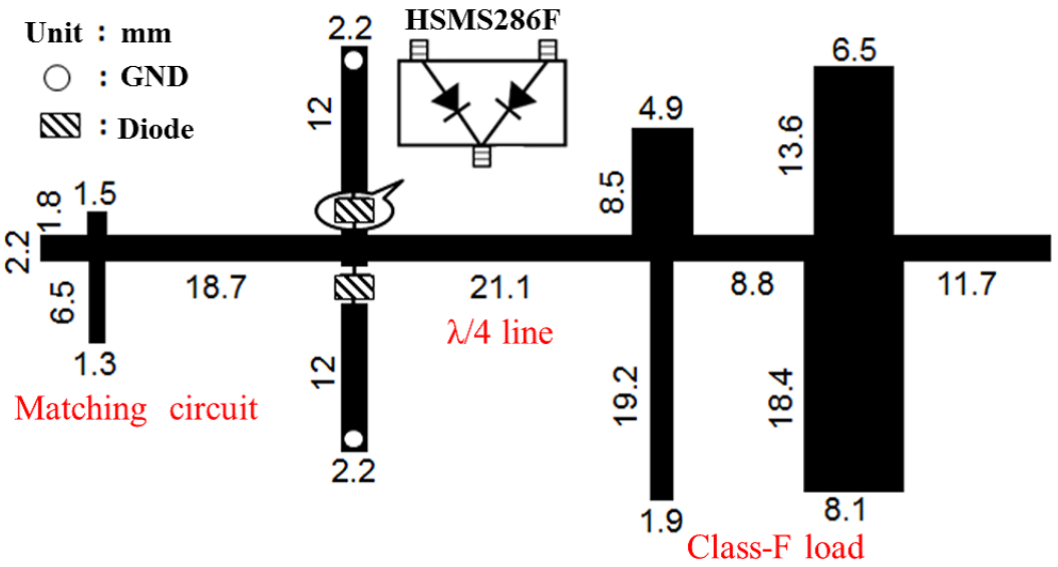
Table 4.1: Efficiency and output voltage with different input powers for a fixed load of 190 Ω.

Input power (mW)	40	60	80	100
Efficiency (%)	80.2	81.9	82.5	81.4
Output voltage (V)	2.46	3.06	3.54	3.93

matching at the input port. Second, the DC blocking capacitor is removed because the equivalent series resistance of the capacitor causes some losses. Instead of a DC blocking capacitor in the rectifying circuit, a DC blocking element is used in the experimental equipment to block DC voltage from diodes. Third, the same rectifying diode of HSMS286F is used but the polarity of the diode is inverted to obtain a positive output voltage as usual. In simulation, we optimize the dimensions of each microstrip line to make a trade-off between reflection, diode loss, high harmonic loss and the efficiency until obtaining a high efficiency. Figure 4.1(a) shows the layout circuit and the size of the designed rectifier. We use the same dielectric substrate of NPC-F260 whose parameters are listed in Table 3.4. Figure 4.1(b) shows the photograph of the fabricated rectifier and the whole size of the circuit is approximately as same as the circuit in Chapter 3.4.

Similarly, the efficiency of the rectifying circuit is confirmed and the results are shown in Fig. 4.2. Figure 4.2(a) shows the experimental results of the efficiency versus the load resistance from 100 to 400 Ω for different input powers such as 40, 60, 80, 100 mW. It is obvious that the efficiency is varying with the load resistance. As seen from the figure, the optimal load is approximately steady at 190 Ω even slight smaller as the input power is increasing. At the load of 190 Ω, the peak efficiency and output voltage of each input power level are presented in Table 4.1. In these input power ranges, the peak efficiency is changing from 80.2 % to 82.5 %, where the variation is small. In addition, For a higher input power, the efficiency curve rises early and also drops rapidly with the increase of load resistance. This is because the higher input power gets a higher output voltage which causes the voltage applied on the diode exceeding the breakdown voltage.

Figure 4.2(b) shows the simulation and experimental results of the efficiency versus the input power varying from 20 to 120 mW when the load resistance is



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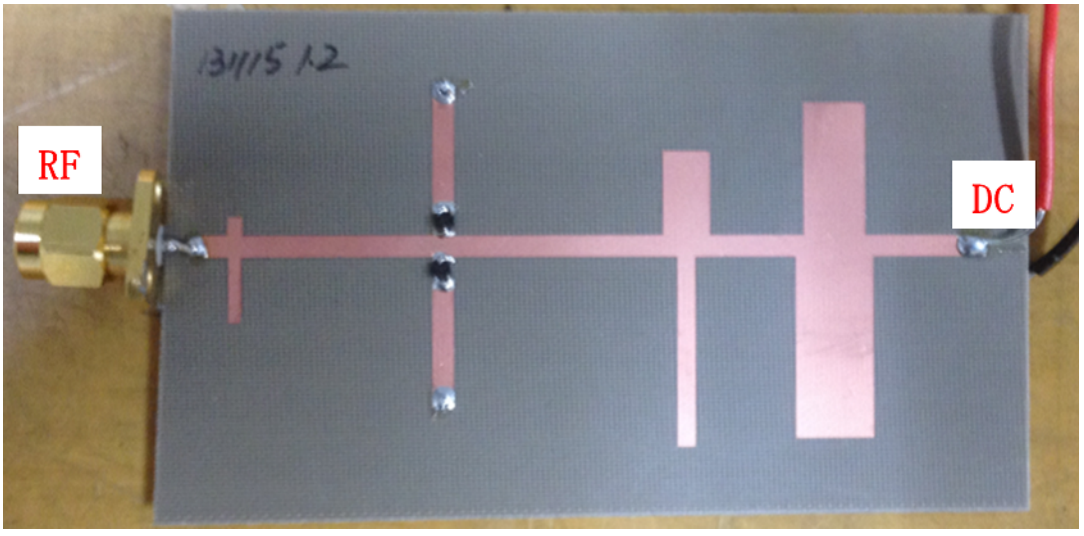

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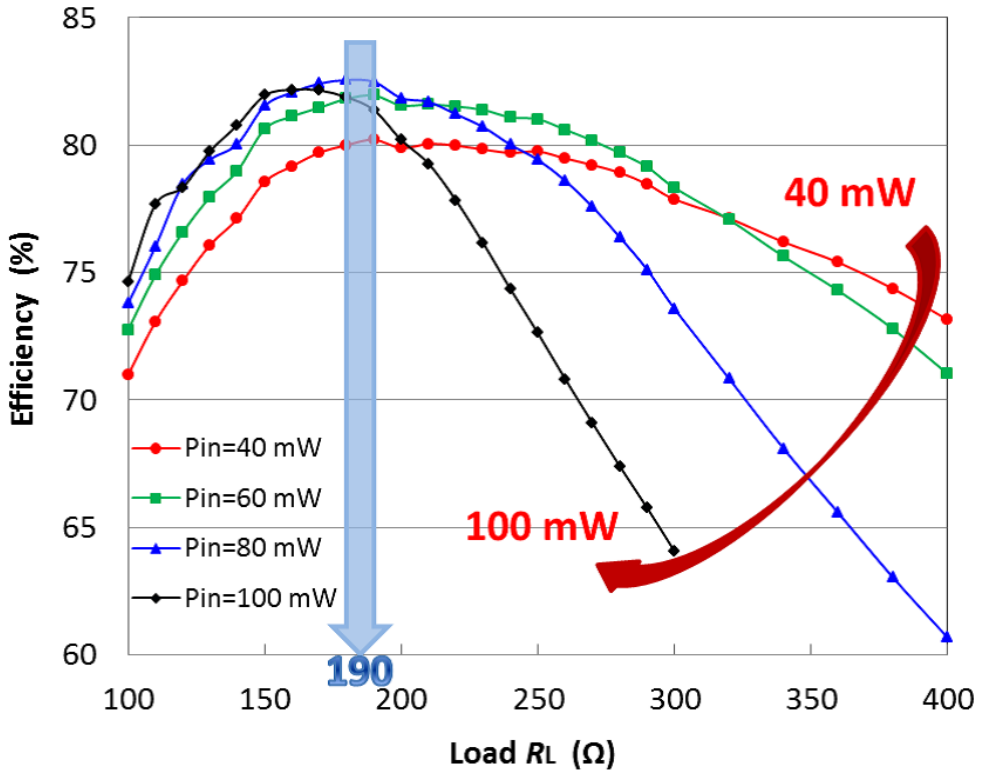
Figure 4.1: Proposed positive output voltage rectifier. (a) Structure and size. (b) Photograph.



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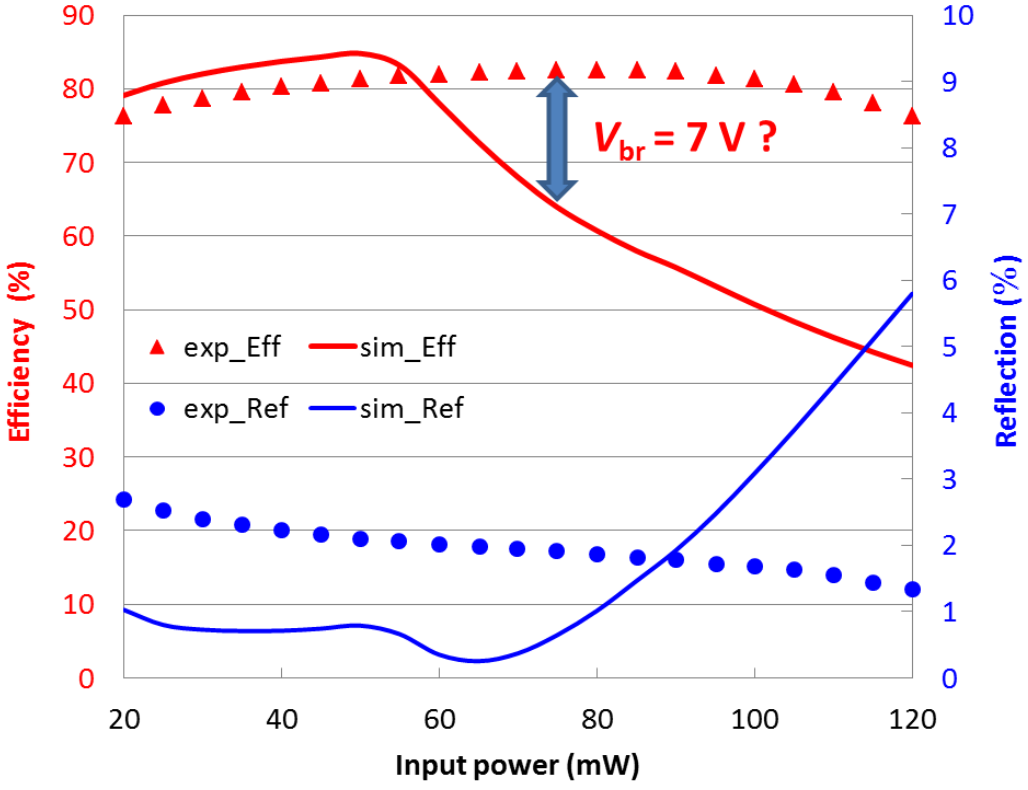


(b)

Figure 4.2: Simulation and experimental results of the high-efficiency rectifier. (a) Efficiency vs. load resistance. (b) Efficiency vs. input power.

fixed at 190 Ω. In the case of the simulation, maximum efficiency is 84.7 % at the input power of 50 mW. In the case of the experiment, maximum efficiency of 82.5 % is obtained at input power of 80 mW. The experiment results almost agree with the simulation when the input power is below 60 mW. However, they are greatly different with the simulation results when the input power is over 60 mW. There are some differences between simulation and experiment results as same as the rectifying circuit in Chapter 3.4. The diode model of di_hp_HSMS286F_20000301, which is a high frequency diode model in ADS library, is used in the simulation. This diode model uses the SPICE parameter where the breakdown voltage B_{V} is 7 V. The reverse current will reach the limitation value of I_{BV} when the reverse voltage reaches B_{V}, and it will increase when the reverse voltage exceeds the B_{V} until the reverse direction of the diode is completely conducted. In simulation, the impedance of the diode is greatly changed when the voltage applied on the diode exceeds the breakdown voltage. Hence, the reflection and diode losses increase rapidly which leads to a decrease in efficiency. However, in experiment, the real breakdown voltage of the diode is far over 7 V that the efficiency falls slowly as the diode loss is increasing even though the power exceeds the power value at peak efficiency point. There is an actual example that a diode with breakdown voltage of 9 V in SPICE parameter can bear a reverse voltage over 20 V in experiment [70].

4.3 Self-powered DCM Buck-boost Converter

4.3.1 Self-powered Non-inverting Buck-boost Converter

Compared to the externally-powered buck-boost converter in Chapter 3, this chapter proposes a self-powered non-inverting buck-boost converter as shown in Fig. 4.3. The main circuit also consists of capacitors, an inductor, an N-MOS and a diode whereas the diode and GND position are changed. Due to those changes, the converter affords a positive output voltage with a positive input voltage. The control-pulse circuit only includes two oscillators which are powered by the input voltage of the converter. Similarly, the operating waveform is as same as Fig. 3.4(b) while the current flowing direction is different. There are three state intervals when the LF oscillator is at work-state. The on-state is that the current flows through the inductor and N-MOS to the GND when the gate voltage is at

high-level. The off-state is that the current flows through the inductor and diode to the load resistance when the gate voltage is at low-level. The zero-state is that no current flows through the circuit which is as same as the sleep-state of the LF oscillator.

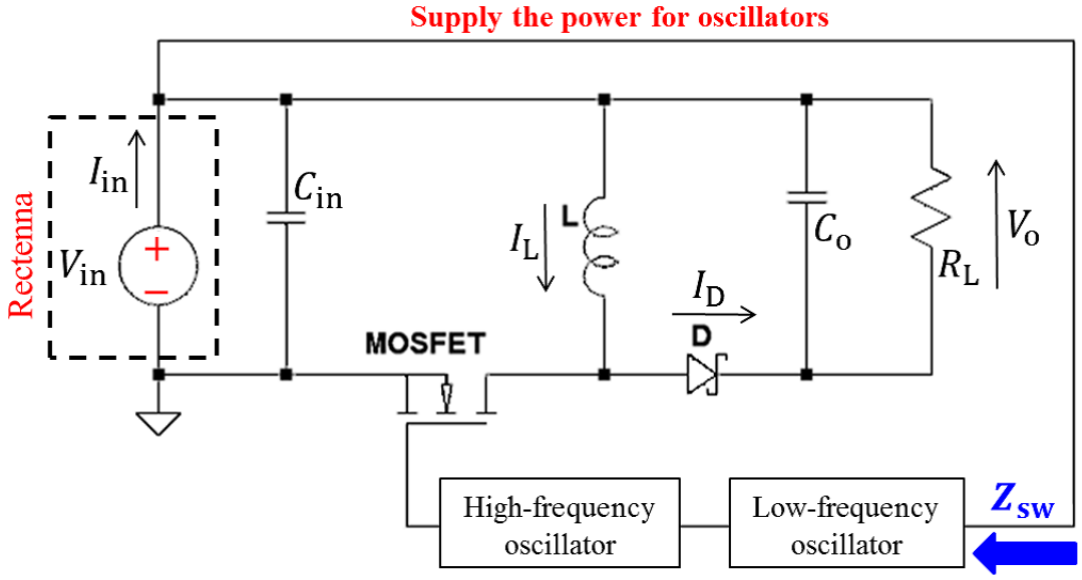


Figure 4.3: Block diagram of the self-powered non-inverting buck-boost converter.

Therefore, this buck-boost converter also operates at discontinuous conduction mode. Assuming R'_{in} to be input resistance of this converter, it is the combination of R_{in} (Eq. (3.37) mentioned in Chapter 3.3) and the impedance of the control-pulse circuit Z_{sw} in parallel as seen in Fig. 4.3. Hence, R'_{in} satisfies the following equation:

$$R'_{in} = \frac{R_{in}}{1 + \frac{R_{in}}{Z_{sw}}}. \tag{4.1}$$

To measure the impedance of the control-pulse circuit Z_{sw}, the simulation and experiment on the control-pulse circuit are conducted and Fig. 4.4 shows the schematic of the simulation circuit and the simulation waveform in LTspice. As same as the previous converter, LF and HF oscillators utilize LTC6992 and LTC6900, respectively. Figure 4.4(b) shows the operating waveforms such as LF pulse V_ι and the control pulse V_{gate} at the input voltage of 3 V. Figure 4.5 shows the measured waveforms of the control-pulse circuit at the supplied voltage of 3 V, where the high, low frequency are 10 kHz and 98 Hz, respectively. The amplitude of V_h and V_ι are both 3.96 V, which is almost equal to the supplied voltage of 3 V.

Figure 4.6(a) shows the simulation and experimental results of the impedance Z_{sw} which indicates that Z_{sw} is large enough to exceed 8 kΩ when the supplied voltage is over 3 V. This value also agrees with the calculated value of 8.5-14 kΩ, according to the data sheet of LTC6992 [71]. By the way, the power dissipation of the control-pulse circuit is also investigated as shown in Fig. 4.6(b), which indicates that the power consumption is small and increasing as the supplied voltage increases.

Practically, the input resistance of the buck-boost converter $R_{\mathrm{in}}^{'}$ is designed close to 190 Ω which is the optimal load of the rectifying circuit in RF-DC-DC circuit. Compared with Z_{sw}, 190 Ω is small enough, which leads to following:

$$R_{\mathrm{in}}^{'} \approx R_{\mathrm{in}}. \tag{4.2}$$

Therefore, Eq. (3.37) is also adapted to $R_{\mathrm{in}}^{'}$.

4.3.2 Parameters Selection

According to Eq. (3.37), the R_{in} is decided by the inductance, two pulse duty-on ratios and the high frequency. Furthermore, the designed rectifying circuit exhibits a peak efficiency at the load of 190 Ω. Hence, it is necessary to make R_{in} equal to 190 Ω with suitable value of four parameters. Figure 4.7 shows the schematic diagram of the HF oscillator LTC6900 and LF oscillator LTC6992. LTC6900 is a precision, low power oscillator that is easy-to-use and occupies very little PC board space. The oscillator frequency is programmed by a single external resistor (R_{SET}) as shown in Fig. 4.7(a). The three-state DIV input determines whether the master clock is divided by 1, 10 or 100 before driving the output, providing three frequency ranges spanning 1 kHz to 20 MHz. The oscillator can be easily programmed using the simple formula below [72]:

$$f_{\mathrm{h}} = \frac{10\mathrm{MHz}}{N}\frac{20\mathrm{k\Omega}}{R_{\mathrm{SET}}}, N = 1, 10, 100. \tag{4.3}$$

If we supply a 3 V voltage to pin 4 and assume R_{SET} as 200 kΩ, then, N becomes 100 and f_{h} can be calculated to be 10 kHz referring to Eq. (4.3). In

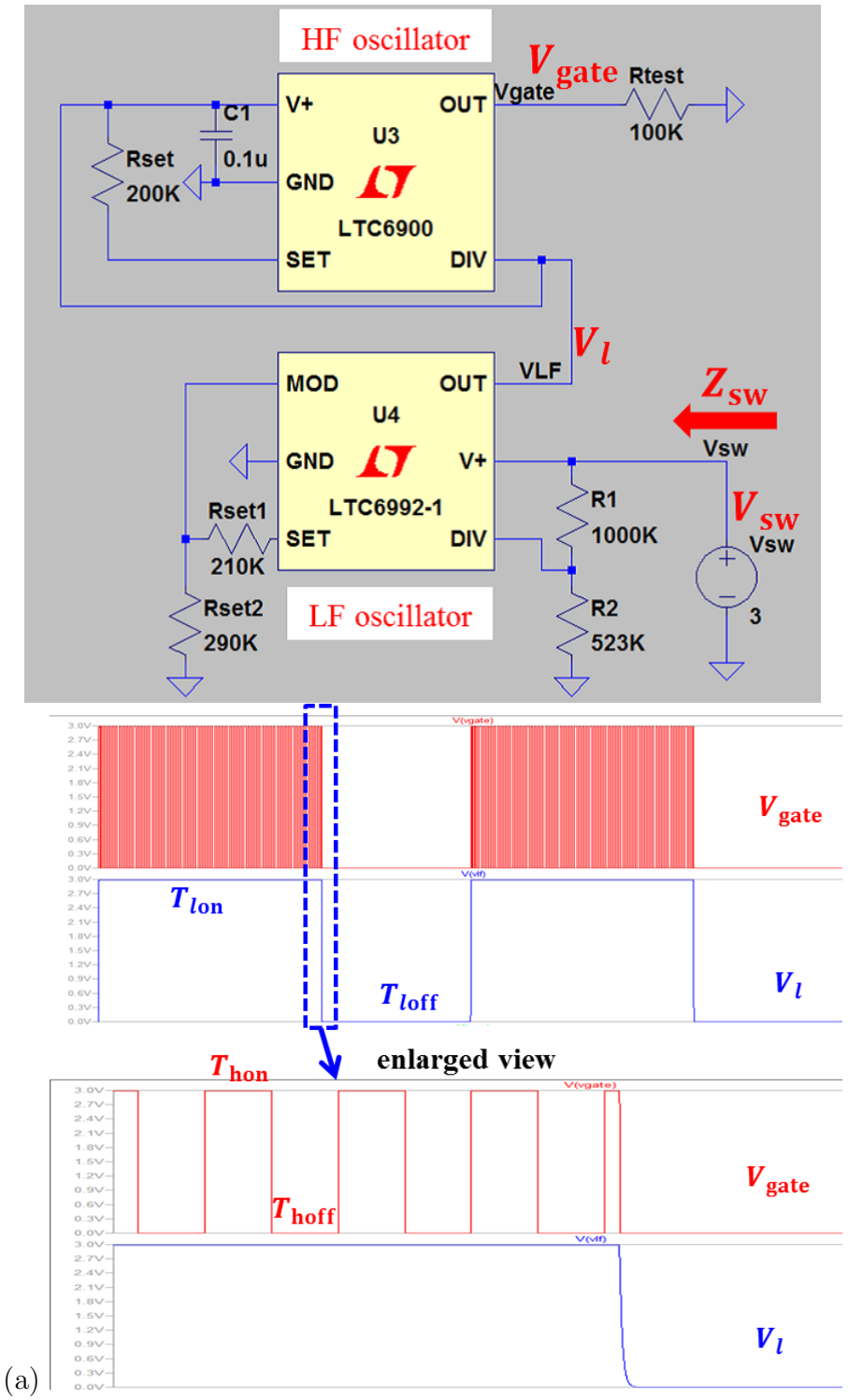


Figure 4.4: Control-pulse circuit of the buck-boost converter in simulation (LTspice). (a) Simulation circuit: LTC6992 is the LF oscillator; LTC6900 is the HF oscillator. (b) Simulated waveforms.

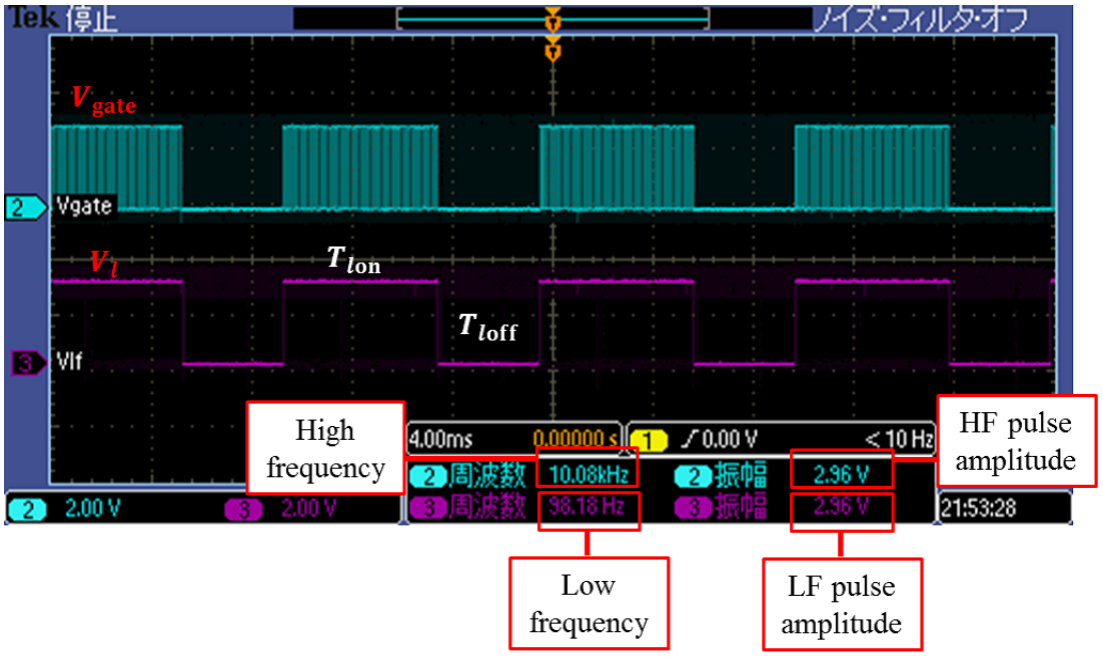


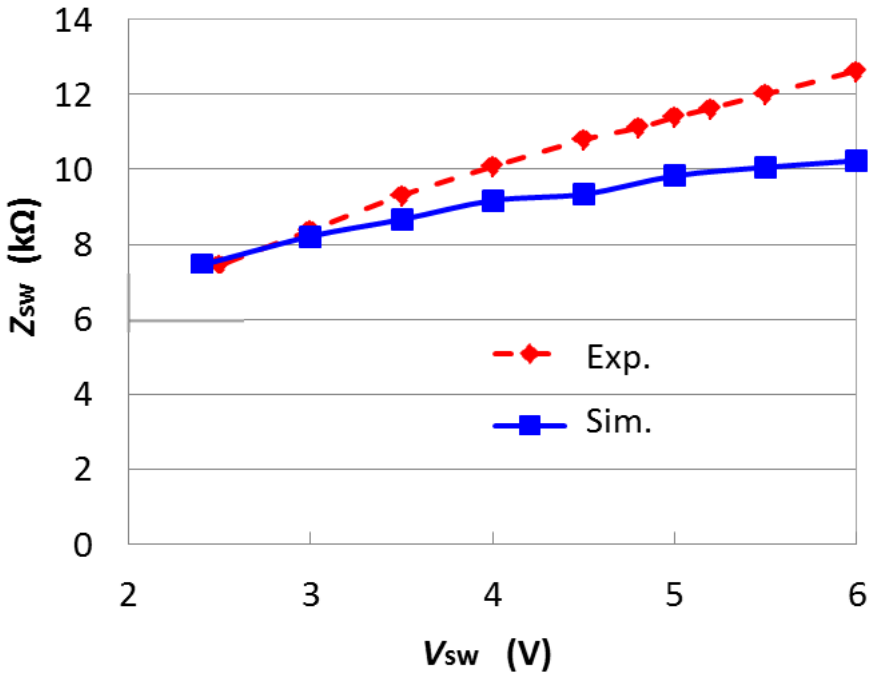
Figure 4.5: Measured waveforms of the control-pulse circuit.

addition, LTC6900 operates from a single 2.7 to 5.5 V supply and outputs a square pulse with the duty-on ratio of 0.5. Hence, D_{h} is 0.5.

The LF oscillator LTC6992 is a silicon oscillator with an easy-to-use analog voltage-controlled pulse width modulation (PWM) capability. The resistance, R_{SET1} and R_{SET2}, program the LTC6992's internal master oscillator frequency. The output frequency is determined by this master oscillator and an internal frequency divider (N_{DIV}), programmable to eight settings from 1 to 16384. The divider ratio N_{DIV} that relates to R_1 and R_2 is set by a resistor divider attached to the DIV pin. With a single 2.25 to 5.5 V supply, the output frequency f_ι, exhibits wide from 3.81 Hz to 1 MHz, can be expressed as following [71]:

$$f_\iota = \frac{1\mathrm{MHz}}{N_{\mathrm{DIV}}} \frac{50\mathrm{k}\Omega}{R_{\mathrm{SET1}} + R_{\mathrm{SET2}}}, N_{\mathrm{DIV}} = 1, 4, 16....16384. \tag{4.4}$$

Here, the divider ratio N_{DIV} and $R_{\mathrm{SET1}} + R_{\mathrm{SET1}}$ are chosen to be 1024 and 530 kΩ, respectively. According to Eq. (4.4), f_ι becomes approximately 100 Hz. The LF f_ι does not effect R_{in} as seen from Eq. (3.37). Whereas the efficiency of the converter will decrease because switching loss is increased as f_ι rises. On the other hand, if f_ι is too low, gate-off time of the MOSFET gets longer which results in a big output capacitor to stabilize power supply to the load. Therefore, f_ι of 100 Hz is selected.



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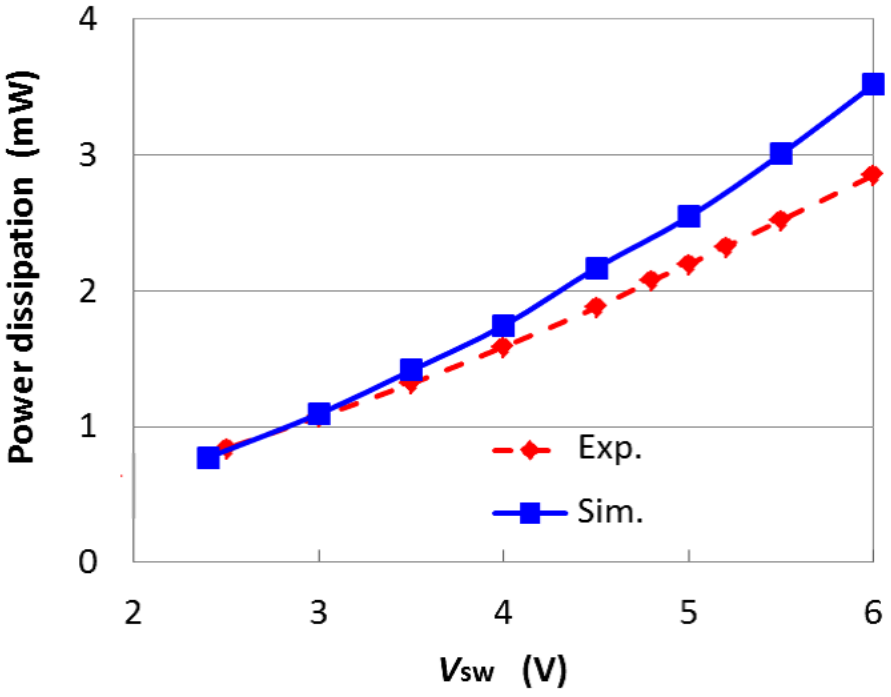

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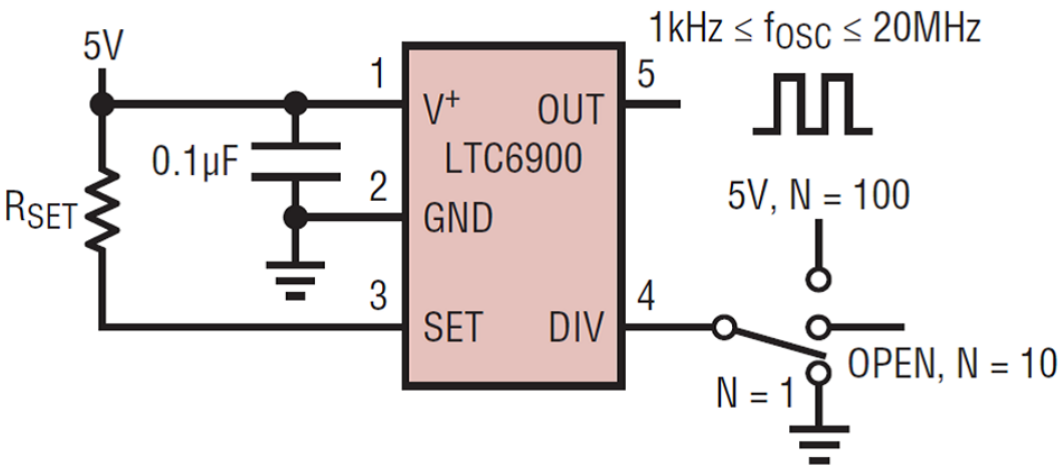
Figure 4.6: Simulation and experimental results of the control-pulse circuit. (a) Impedance Z_{sw}. (b) Power consumption.



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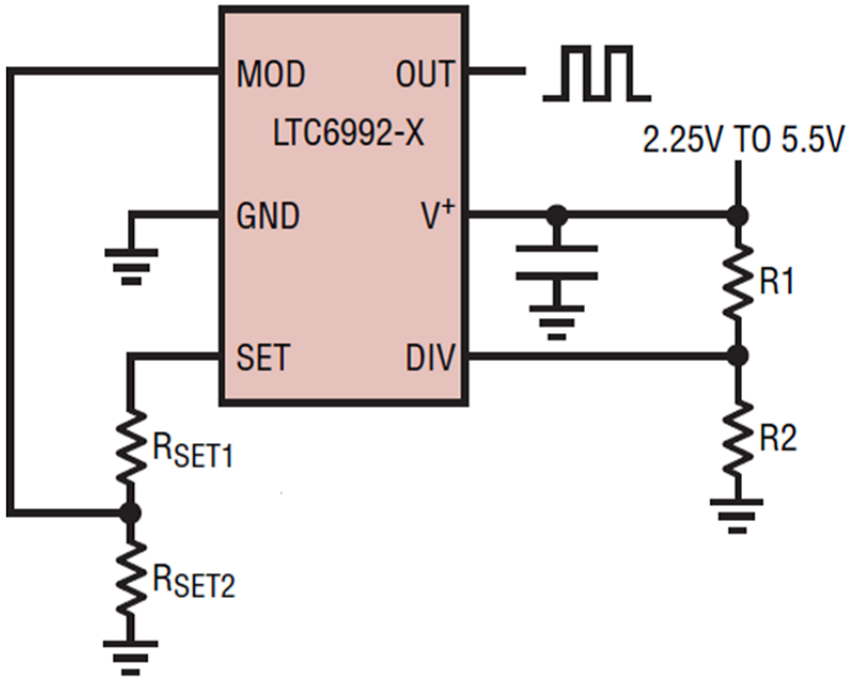


(b)

Figure 4.7: Block diagram of oscillators. (a) HF oscillator LTC6900 [72]. (b) LF oscillator LTC6992 [71].

Figure 4.7(b) shows a simple circuit for generating an arbitrary duty cycle at a fixed frequency. The LF duty-on ratio D_ι can operate wide from 0 % to 100 % and it can be expressed as below [71]:

$$D_\iota = \frac{5}{4}\frac{R_{\mathrm{SET2}}}{R_{\mathrm{SET1}} + R_{\mathrm{SET2}}} - \frac{1}{8}. \tag{4.5}$$

We use a variable resistance of 530 kΩ which is divided into two parts such as R_{SET1} and R_{SET2}. We assign R_{SET2} to be 234 kΩ, then, D_ι becomes approximately 0.427. Here, a variable resistor is used, then the proportion of R_{SET2} and R_{SET2} is adjustable to change D_ι. Consequently, the R_{in} is changeable. So far, the four parameters are decided as following: $f_{\mathrm{h}} = 10$ kHz, $D_{\mathrm{h}} = 0.5$, $D_\iota = 0.427$, and inductance L is chosen to be 1 mH. Applying these parameters into Eq. (3.37), R_{in} becomes 187 Ω, which is very close to the optimal load (190 Ω) of the rectifying circuit. It means that the input resistance of the new buck-boost converter is designed to equal the optimal load of the rectifying circuit in theoretical analysis.

With respect to the other components of the converter such as diode, N-MOS, inductor and capacitors are the same as the one outlined in Table 3.3, while the input capacitor $C_{\mathrm{in}} = 220$ μF ($V_{\mathrm{rated}} = 35$ V) is selected. The ripple of the voltage is small if the input voltage supplied from a DC power source. But the input voltage of the converter utilizes the output voltage of the rectifying circuit where the ripple is very big. In order to limit the input voltage ripple of the converter in RF-DC-DC circuit, a relatively high input capacitance is selected. Due to input voltage ripple, the efficiency varying ripple is much big in Figs. 3.14(a) and 3.15. Additionally, the MOSFET gate driver is removed in this new buck-boost converter. A MOSFET gate driver is usually applied to short switching time to minimize the switching loss for a large gate capacitance MOSFET. However, the N-MOS FDS6298 has been designed specifically to improve the overall efficiency of DC-DC converters using either synchronous or conventional switching PWM controllers. It has been optimized for low gate charge, low $R_{\mathrm{DC(on)}}$ and fast switching speed [73]. Note that the output of the HF oscillator directly drives the gate of the N-MOS in Fig. 4.3. Compared with the control-pulse circuit of the previous converter, the new one will dissipate less power because of no power consumption from gate diver.

4.3.3 Input Resistance Verification

With surface-mount device (SMD) resistors instead of axial-lead resistors, the new buck-boost converter mounts on a 3 cm × 3 cm PCB as shown in Fig. 4.8. Then, to verify the input resistance, simulate the circuit using LTspice and conduct an experiment using the measurement circuit as shown in Fig. 4.9(a) where does not need an extra DC power source. Figure 4.9(b) shows the operating waveform of V_{gate} at the input voltage of 3 V. Observe that the high frequency is 10 kHz, and the amplitude is 2.88 V which is a little smaller than the one measured in control-pulse circuit in Chapter 4.3.2. This is because the impedance of gate-source is not big enough to neglect the output impedance of the HF oscillator resulting in a drop of V_{gate}. However, this voltage drop is sufficiently small that does not affect driving the gate of the N-MOS.

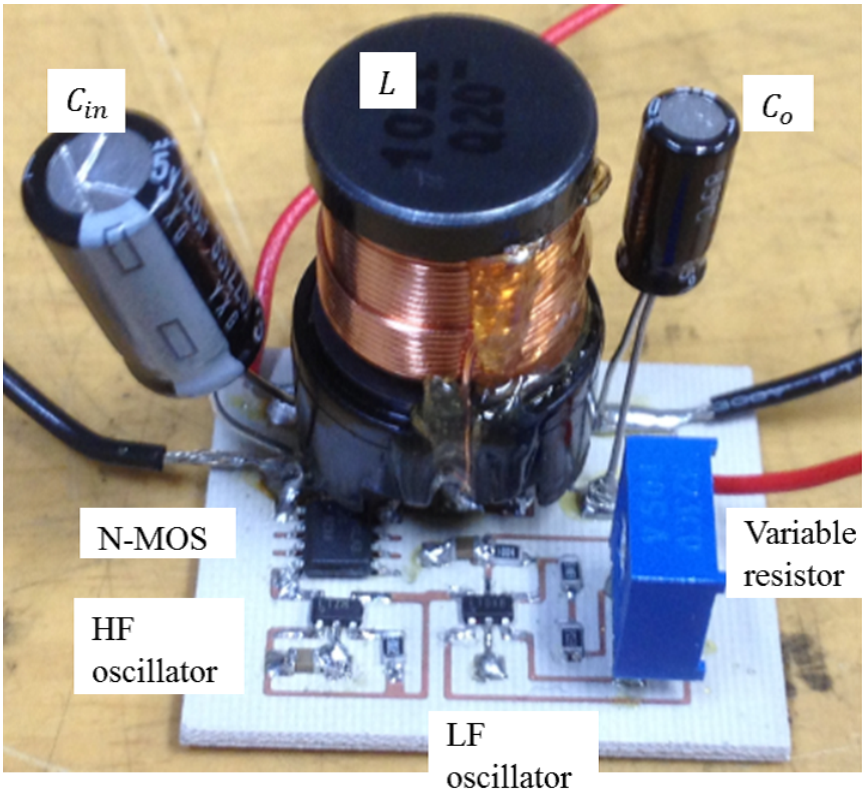
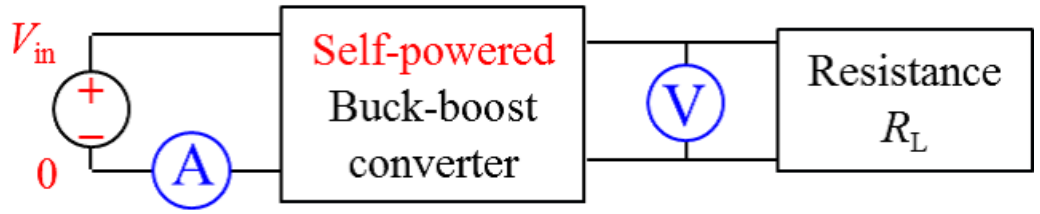


Figure 4.8: Photograph of the self-powered non-inverting buck-boost converter.

The simulation and experimental results of the self-powered buck-boost converter are shown in Fig. 4.10, where Sim_Eff and Exp_Eff represent the efficiency of the simulation and experimental, Sim_Rin and Exp_Rin represent the input resistance of the simulation and experimental, respectively. Figure 4.10(a), at a fixed input voltage of 3.5 V, shows the efficiency and the input resistance R'_{in}



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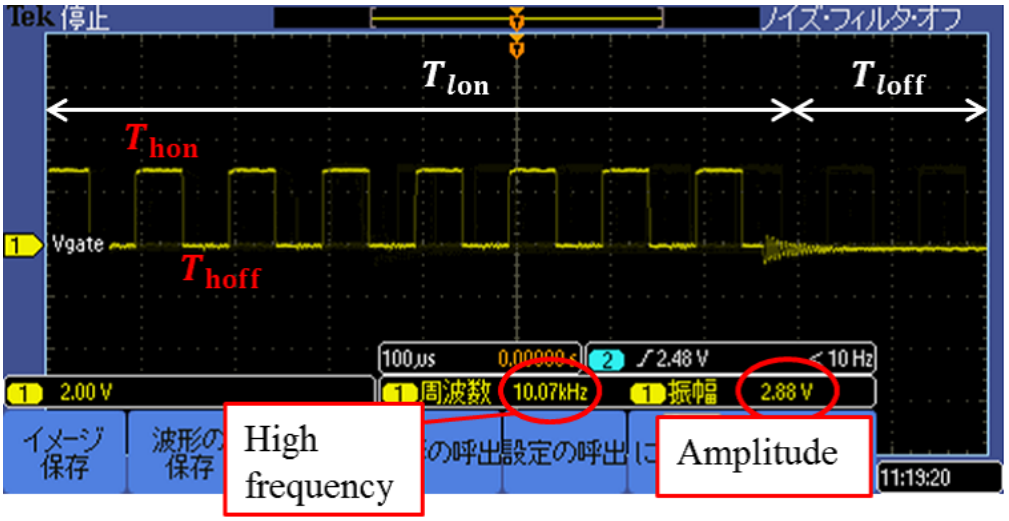


(b)

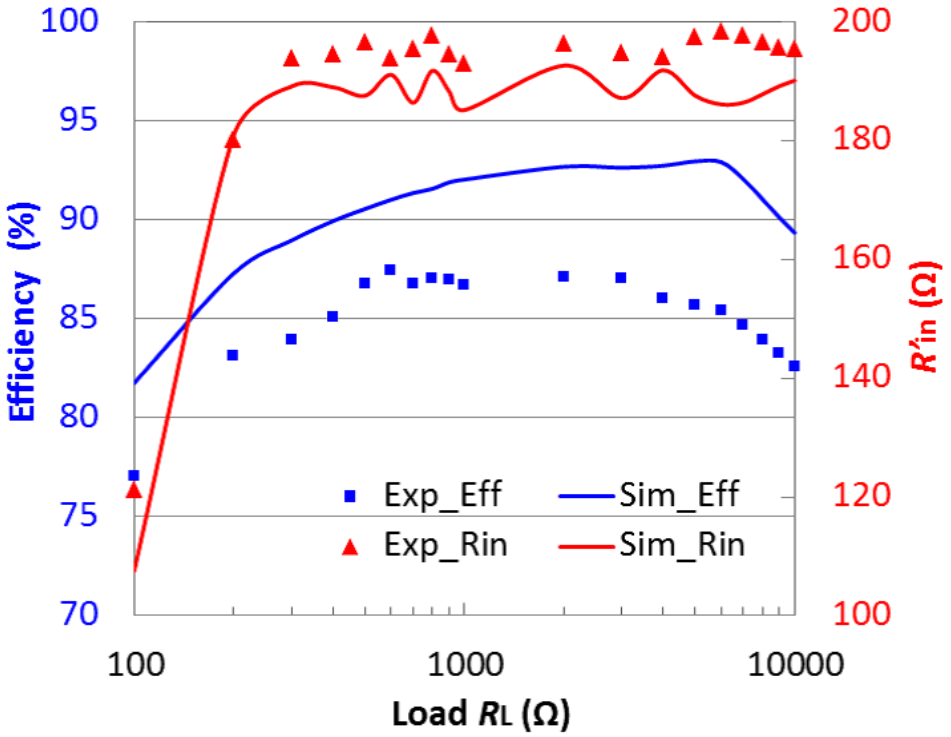
Figure 4.9: Experiment setup and measured waveform of the converter. (a) Experiment setup. (b) Measured waveform of V_{gate}.

versus the load changing from 100 Ω to 10 kΩ. Here, the input voltage of 3.5 V is selected to agree with the output voltage of the rectifying circuit when it performs maximum efficiency at the optimal load for an RF input power of 80 mW. Furthermore, the input power of the converter calculated to be 63 mW which almost agrees with the output DC power of the rectifying circuit calculated from Table 4.1. In the case of R'_{in}, the simulation and experimental results are in well agreement with the designed value of 187 Ω when the load is wide changing from 200 Ω to 10 kΩ. At the load of 100 Ω, when the load resistance is smaller than the input resistance, the buck-boost converter operates at the CCM which leads to the input current increasing. As a result, the input resistance is decreased. In the case of the efficiency, it increases at low load resistance and decreases at the high load resistance. The output current is decreasing as the load increases at a fixed input power. The major loss of the converter is dominated by the loss such

as diode loss, switching loss, inductor and other resistance losses when the load resistance is below 3000 Ω. Whereas the major loss of the converter is dominated by the loss of the output capacitor that a part of the output current flows through the output capacitor as the load resistance is too big to neglect the impedance of the capacitor. The measured efficiency is over 85 % in the load range from 400 to 6000 Ω.

Similarly, Figure 4.10(b), at a fixed load resistance of 800 Ω, shows the efficiency and the input resistance $R_{\text{in}}^{'}$ versus the input voltage changing from 2.5 to 4.5 V. At this input voltage range, the input power is varying from 31 to 104 mW. Note that the $R_{\text{in}}^{'}$ is also almost steady at 190 Ω and is independent of the input voltage. As for the efficiency, the measured one is over 85 % at the whole input voltage range. In the case of the simulation, the efficiency is increasing as the input voltage increases. In the case of the experiment, the efficiency is increasing at the beginning and then slightly decreases as the input voltage increases. Because of the limitation from the operation voltage of HF and LF oscillators, the input voltage is restrained among the range of 2.5 to 6 V. The switching loss increases because of a low gate control voltage V_{gate} when the input voltage is low. The diode loss, inductor and other resistance loss will increase at a high input voltage due to current increasing. Additionally, as seen from Fig. 4.6(b), the power consumption of oscillators is increasing as the supplied voltage increases.

In addition, for both figures in Fig. 4.10, the measured $R_{\text{in}}^{'}$ is slightly bigger than the simulation results. The reason is that the wire resistance and the noise of real pulse wave cuts the on-state time which minimizes the input current. Furthermore, the experiment efficiency is approximately 5 point lower than the simulation one because the series resistance loss in capacitor and the wire resistance loss are not considered in simulation. In Chapter 4.3.1, we have discussed that the input resistance of this new buck-boost converter $R_{\text{in}}^{'}$ is the combination of the main circuit R_{in} and the control-pulse circuit Z_{sw} in parallel. In practice, using measurement circuit in Fig. 4.9(a), the input resistance, calculated from measured input voltage and input current, is the value of $R_{\text{in}}^{'}$ which already includes R_{in} and Z_{sw}. According to the simulation and experiment results reported above, the input resistance $R_{\text{in}}^{'}$ is almost equal to the value of R_{in}, which well agrees with the theory analysis. Consequently, that the new buck-boost converter has constant input resistance characteristic is verified both by simulation and experiment results.



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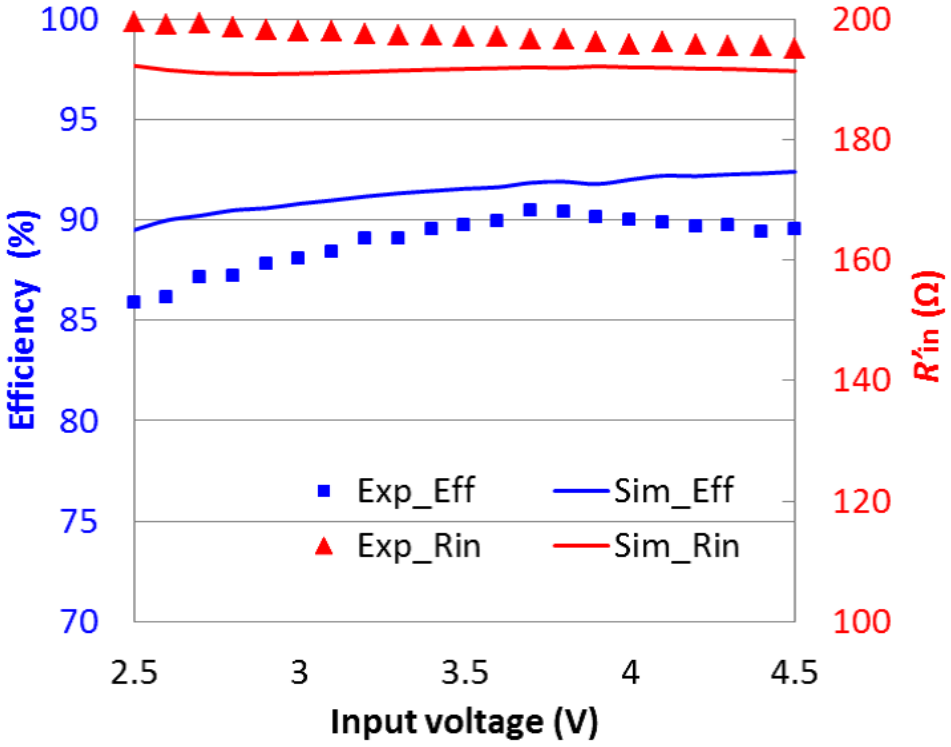
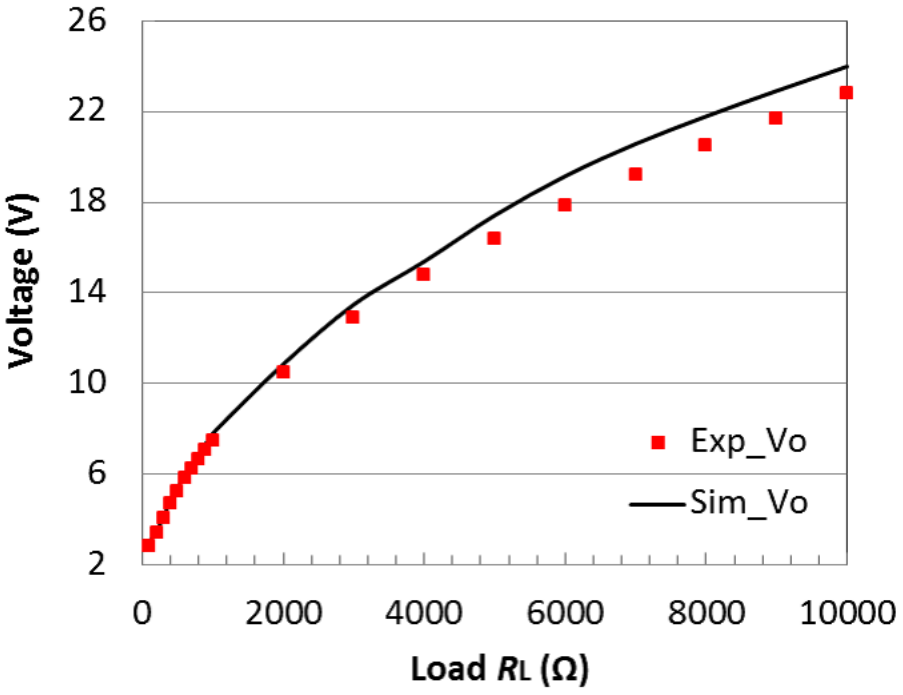
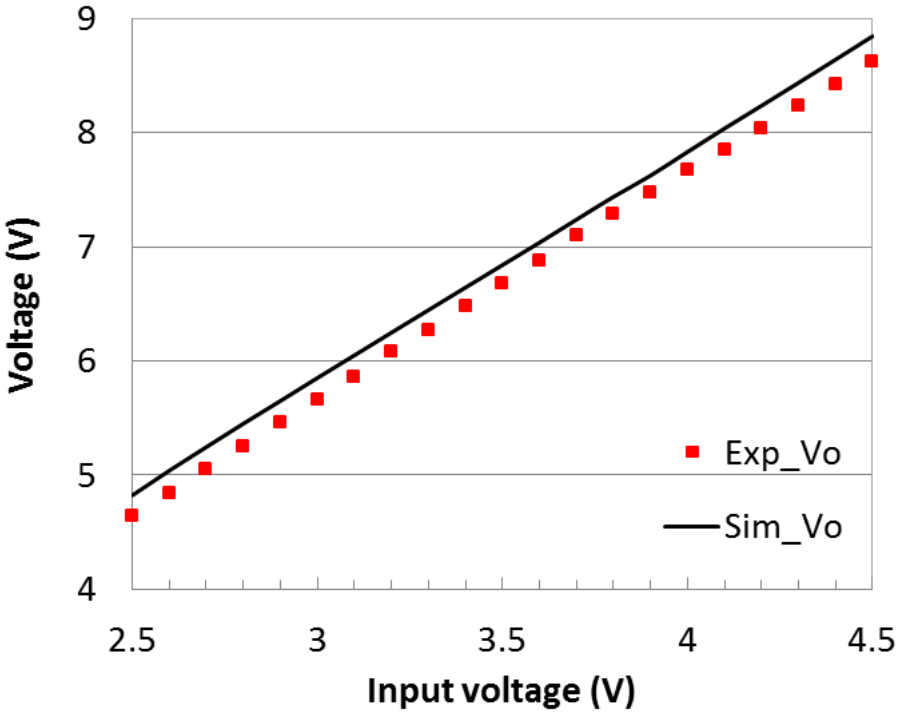


(b)

Figure 4.10: Simulation and experimental results of the self-powered buck-boost converter. (a) Efficiency and input resistance R'_{in} vs. load resistance for a fixed input voltage of 3.5 V. (b) Efficiency and input resistance R'_{in} vs. input voltage for a fixed load of 800 Ω.



(a)



(b)

Figure 4.11: Simulation and experimental results about output voltage of the self-powered buck-boost converter. (a) Output voltage vs. load resistance for a fixed input voltage of 3.5 V. (b) Output voltage vs. input voltage for a fixed load of 800 Ω. Exp_V_o, Sim_V_o are the experiment result and simulation result, respectively.

With respect to the input/output voltage relationship of this buck-boost converter, apply the formula of $\frac{V_o^2}{R_L} = \frac{V_{in}^2}{R_{in}}$ to Eq. (3.37), it yields:

$$V_o = \sqrt{\frac{D_h^2 D_\iota}{2Lf_h} V_{in}^2} \cdot R_L^{\frac{1}{2}}, \tag{4.6}$$

$$V_o = \sqrt{\frac{D_h^2 D_\iota}{2Lf_h} R_L} \cdot V_{in}, \tag{4.7}$$

which indicates that the relationship between V_o and R_L is a parabolic curve with horizontal axis when the input voltage V_{in} is fixed. Similarly, the relationship between V_o and V_{in} is linear when the load resistance R_L is fixed. Figure 4.11 gives the simulation and experiment results about output voltage of the buck-boost converter. Figure 4.11(a) shows the output voltage versus load resistance for a fixed input voltage of 3.5 V where the output voltage is increasing as the load increases, and the shape of the curve well agrees with Eq. (4.6). Figure 4.11(b) shows the output voltage versus input voltage for a fixed load resistance of 800 Ω. Where the output voltage is linearly increasing as the input voltage increases, and the shape of the curve well agrees with Eq. (4.7). For both figures, the experiment results are in good agreement with the simulations. It is clear that the output voltage is boosted by the buck-boost converter, which indicates that this resistance conversion buck-boost converter only plays a booster role here.

4.4 Experiment on Self-powered RF-DC-DC Circuit

Similarly, summarize the efficiency-load relation of the rectifier and buck-boost converter as mentioned above in Table 4.2. As expected, the overall efficiency of the RF-DC-DC circuit should be steady and over 67 % in the load range of 200 Ω to 10 kΩ. Then,we conduct an experiment on the self-powered RF-DC-DC circuit using the experiment architecture in Fig. 4.12(a) with 2.45 GHz microwave power. Here, we can obtain a DC power at the load port with the incident microwave power where an extra DC power source is needless. The measured items are listed in Table 4.3. Using these measured items, the overall efficiency, the efficiency of

Table 4.2: Efficiency and load relationships of the high-efficiency rectifier and self-powered buck-boost converter in experimental results.

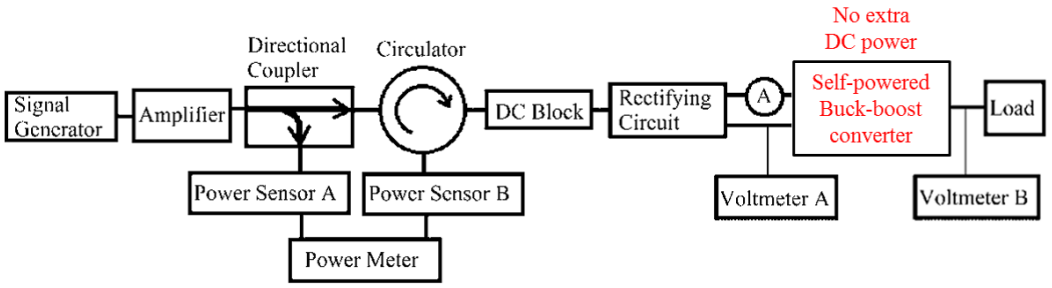
	R_{L}	R_{in}	Efficiency
High-efficiency rectifier	190 Ω		82.5 %
Self-powered buck-boost converter	200 Ω-10 kΩ	almost 190 Ω	over 82 %
Self-powered RF-DC-DC circuit	200 Ω-10 kΩ		predicted value 67 %

Table 4.3: Measured items of the self-powered RF-DC-DC circuit.

Item	Represent
Power sensor A	Input RF power of RF-DC-DC circuit
Power sensor B	Reflection RF power
Voltmeter A	Output voltage of rectifier V_{rec}
Voltmeter B	Output voltage of RF-DC-DC circuit I_{o}
Ammeter	Input current of buck-boost converter I_{in}

the RF-DC rectifier alone, the efficiency of the buck-boost converter alone, the tracking efficiency and the input resistance of the converter can be calculated.

Firstly, several efficiency curves are shown in Fig. 4.13 and each efficiency curve is explained by the block diagram in Fig. 4.14. The red RF-DC-DC curve is the overall efficiency of the RF-DC-DC circuit. The blue RF-DC (with converter) curve is the efficiency of the rectifier when it is connected to the converter. In contrast, the black RF-DC (without converter) curve is the efficiency of a general rectifier. The green DC-DC curve is the efficiency of the buck-boost converter when it is utilized in the RF-DC-DC circuit. With a fixed input RF power of 80 mW, the variation of overall efficiency of the RF-DC-DC circuit over a load range of 100 Ω to 10 kΩ is shown in Fig. 4.13(a). It is obvious that the efficiency of the rectifier (with converter) is almost constant over 80 % in that wide load range. Compared with the general rectifier (black), it seems that the rectifier



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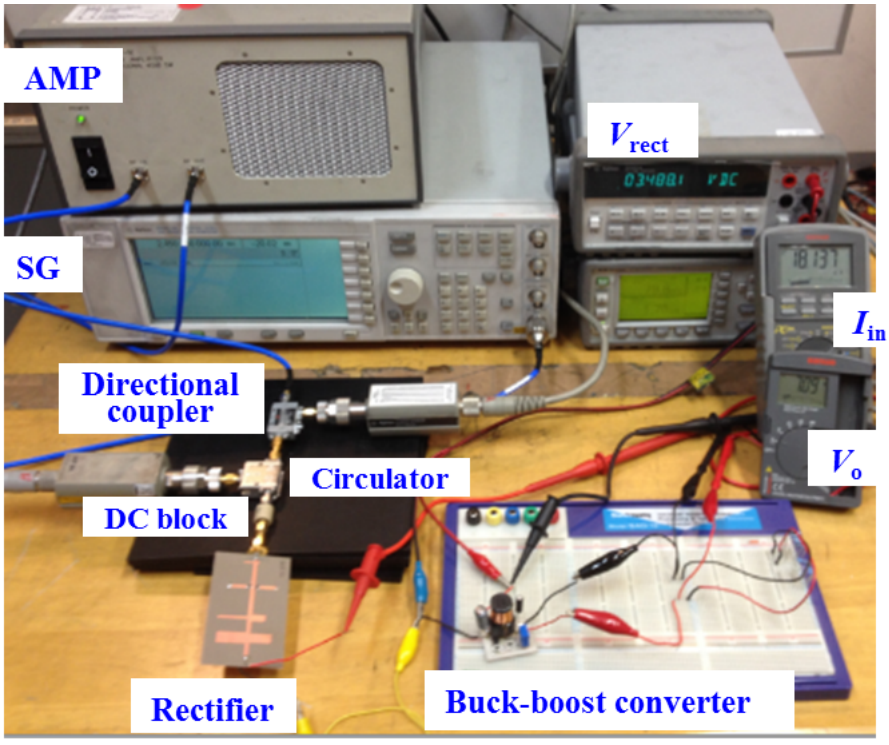


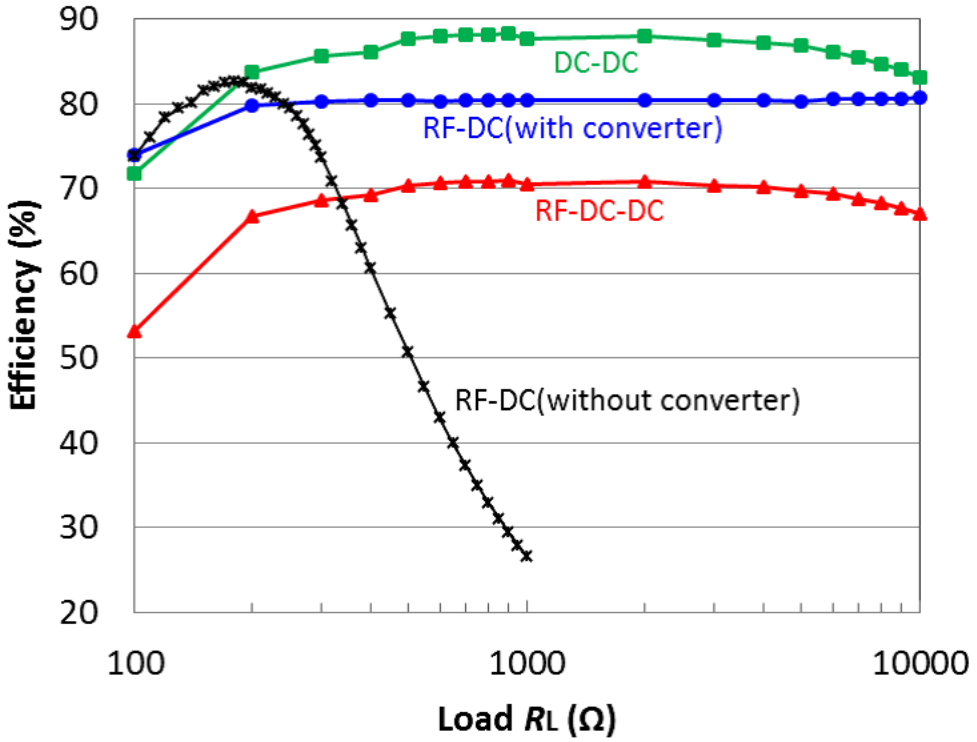
(b)

Figure 4.12: Experiment setup for measuring the efficiency of the self-powered RF-DC-DC circuit. (a) Experiment setup. (b) Experiment photograph.

(with converter) always operates at the peak efficiency despite the load resistance. The efficiency of the buck-boost converter (blue) is over 82 % in that wide load range. Consequently, the overall efficiency, which is the multiplication of previous two efficiencies, is almost constant over 66 % in this ultra-wide load range of 200 Ω to 10 kΩ (a ratio of 1:50). Specially, the overall efficiency is over 70 % in the load range of 400 to 4000 Ω (a ratio of 1:10). The overall efficiency is low at the load of 100 Ω because the input resistance of the converter is mismatching with the optimal load of the rectifier which reduces the rectifying efficiency. As the efficiency of the converter decreases at high load resistance as mentioned in Chapter 4.3.3, the overall efficiency is also slightly decreasing as the load resistance increases.

With a fixed load resistance of 800 Ω, Figure 4.13(b) shows the overall efficiency versus the input power changing from 20 to 120 mW. Note that the overall efficiency exceeds 66 % in the input power range of 40 to 120 mW. The overall efficiency is low when the input power is below 40 mW. Referring to Chapter 4.3.3, the output voltage of the rectifier is smaller than 2.5 V when the input power is lower than 40 mW at the load of 190 Ω. However, a necessary minimum operation voltage for control-pulse oscillators is also 2.5 V. Therefore, the buck-boost converter can not work well when the input power of the RF-DC-DC circuit is below 40 mW, which leads to a decrease in overall efficiency. When the input power exceeds 100 mW, the optimal load of the rectifier becomes lower than the designed value of R'_{in}. As a result, the impedance mismatching occurs between the rectifier and converter. Then the overall efficiency is decreasing as the efficiency of the rectifier decreases. In practice, if the input power is much higher, the output voltage of the rectifier probably reaches the overvoltage (6 V) of control-pulse oscillators which would decrease the efficiency of the converter even break the converter. However, the output DC voltage (V_{DC}) of the rectifier is restrained by the breakdown voltage (B_{V}) of the HF diode which satisfies the equation: $V_{\text{DC}} \approx B_{\text{V}}/2$. As the breakdown voltage of the HF diode is 7 V, the rectifying diode would breakdown first before the output voltage reaching 6 V. It means that the rectifying diode of the rectifier plays another important role in overvoltage protection for control-pulse circuits here. Thus, considering the efficiency and overvoltage, this RF-DC-DC circuit is suggested to be used in the power range of 30 to 140 mW.

The tracking efficiency η_{track}, as defined in Eq. (3.38), is also investigated.



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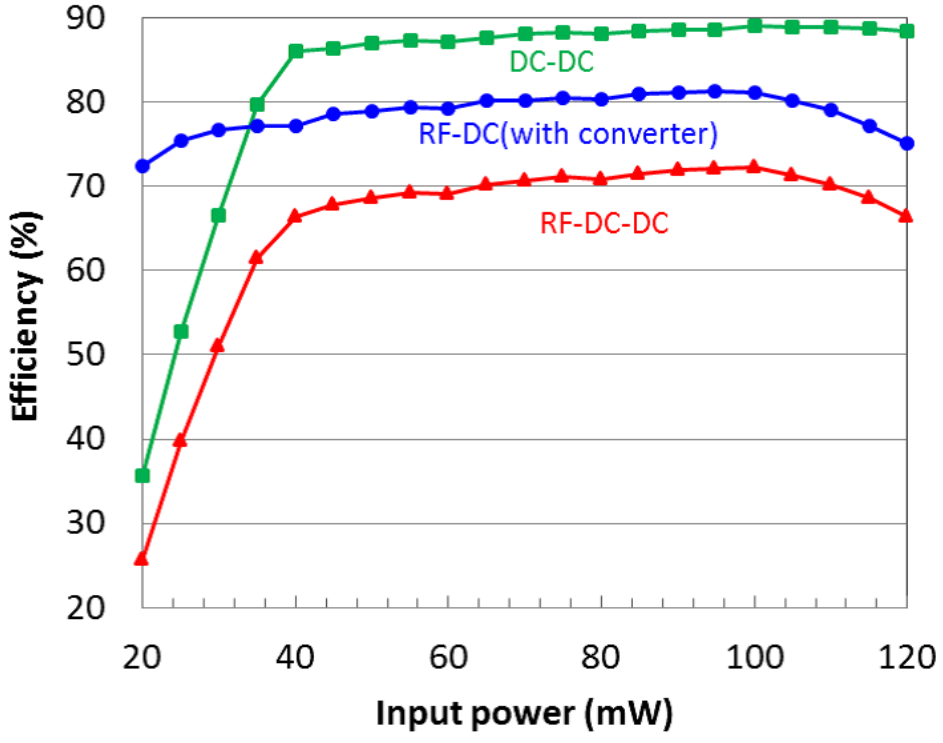
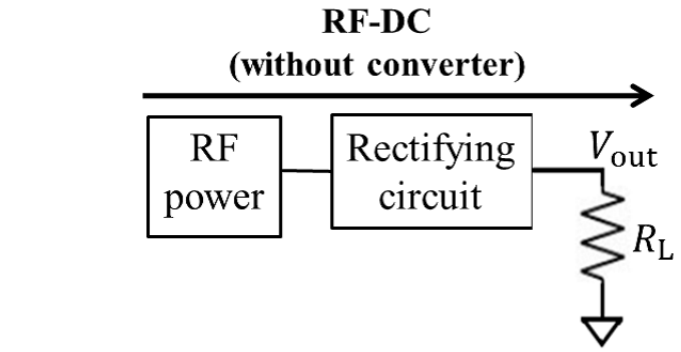


(b)

Figure 4.13: Measured overall efficiency of the self-powered RF-DC-DC circuit. (a) Efficiency comparison of the RF-DC-DC circuit and the general rectifier for a fixed input power of 80 mW. (b) Overall efficiency vs. input power for a fixed load of 800 Ω.



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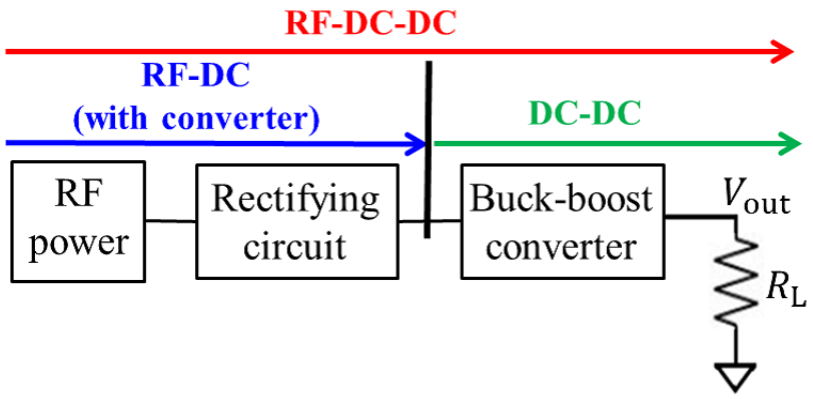


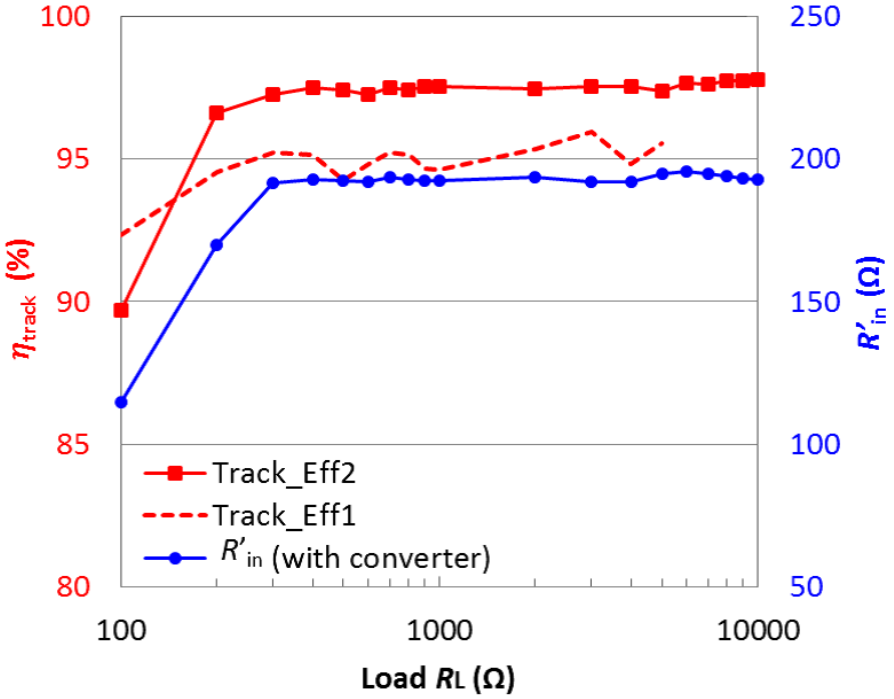
(b)

Figure 4.14: Illustration for the efficiency of each part in RF-DC-DC circuit. (a) General rectifier. (b) RF-DC-DC circuit.

Figure 4.15(a) shows the tracking efficiency and R'_{in} versus load resistance at a fixed input power of 80 mW. Note that the variation tendency of the tracking efficiency is in accord with the input resistance because the former heavily depends on the latter. Track_Eff1 and Track_Eff2 are the tracking efficiency of the externally-powered RF-DC-DC circuit (in Chapter 3.5) and self-powered RF-DC-DC circuit, respectively. It is obvious that Track_Eff2 is constant as high as 97 % in the load range of 200 Ω to 10 kΩ, and is almost 5 percentage points bigger than Track_Eff1 because the impedance matching of the self-powered RF-DC-DC circuit is completely implemented. The R'_{in} is almost constant at the designed value of 190 Ω when the load is over 300 Ω. While it is much smaller than 190 Ω as the load is below 300 Ω and the same phenomenon can be seen in the buck-boost converter from Fig. 4.10(a), where the reason has been explained. Therefore, the overall efficiency is low when the load is below 300 Ω as shown in Fig. 4.13(a).

On the other hand, Figure 4.15(b) shows the tracking efficiency η_{track} and input resistance R'_{in} versus input power at a fixed load resistance of 800 Ω. Similarly, the R'_{in} is almost constant at 190 Ω when the input power is over 40 mW whereas it strictly deviates from 190 Ω when the input power is below 40 mW. In practice, if the input microwave power is too low to obtain the output voltage of 2.5 V in the rectifier, the control-pulse oscillators could not be driven, and the gate voltage of the MOSFET becomes zero which causes the switching to be off. In this case, the input resistance of the buck-boost converter seems to be infinite since the switching is open. Then the output voltage of the rectifier is increasing up to 2.5 V so that it is able to drive the oscillators which leads to switching on. In this case, the input resistance of the buck-boost converter decreases and move to agree with the designed value. After that, the output voltage begins to decrease again until it is too low to drive oscillators, and the operation repeats like above again and again. At this operation condition, the efficiency of the converter is very low. Therefore, R'_{in} is bigger than 190 Ω when the input power is below 40 mW, which is the reason of that the overall efficiency is low at low input power level as shown in Fig. 4.13(b). As for tracking efficiency, Track_Eff2 is also higher than Track_Eff1 and keeps constant over 95 % in the power range of 20 to 100 mW. While it suddenly falls because the optimal load is much smaller than 190 Ω at the input power of 120 mW. This is also the reason of that the overall efficiency is decreased at 120 mW in Fig. 4.13(b).

Figure 4.16 shows the reflection comparison between the rectifier (without converter) and the RF-DC-DC circuit. The reflection of the rectifier heavily depends on the impedance of the rectifying diode which is affected by the output load resistance or the output voltage. In fact, the reflection of the RF-DC-DC circuit is as same as the rectifier, which is also directly related to the output voltage of the rectifier part and the input resistance of the converter. As seen from Fig. 4.16(a), the reflection of the RF-DC-DC circuit is almost constant and is much lower than that of the rectifier (without converter) because the R'_{in} is almost constant at this load range. In Fig. 4.16(b), the reflection of the RF-DC-DC circuit is much higher at low power level because of impedance mismatching as mentioned previously. Additionally, the reflection of the RF-DC-DC circuit is slightly higher (approximate 0.5 percentage point) than that of the rectifier (without converter) because of subtle differences between the actual R'_{in} and 190 Ω.



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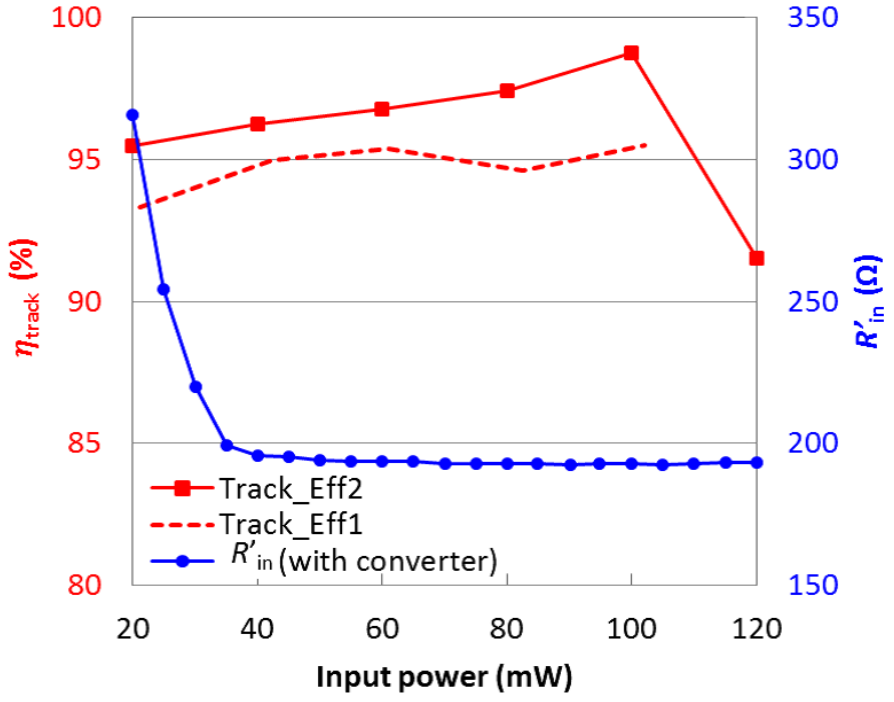
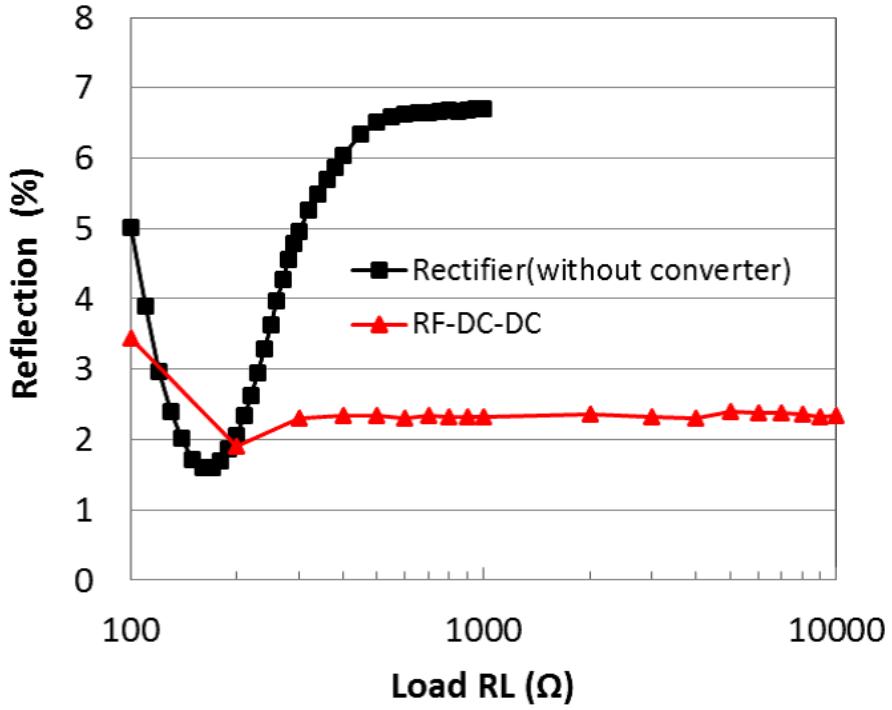


(b)

Figure 4.15: η_{track} comparison and R'_{in}. (a) η_{track} and R'_{in} vs. load resistance for a fixed input power of 80 mW. (b) η_{track} and R'_{in} vs. input power for a fixed load of 800 Ω. Track_Eff1 and Track_Eff2 are the tracking efficiency of the externally-powered and self-powered RF-DC-DC circuit, respectively.



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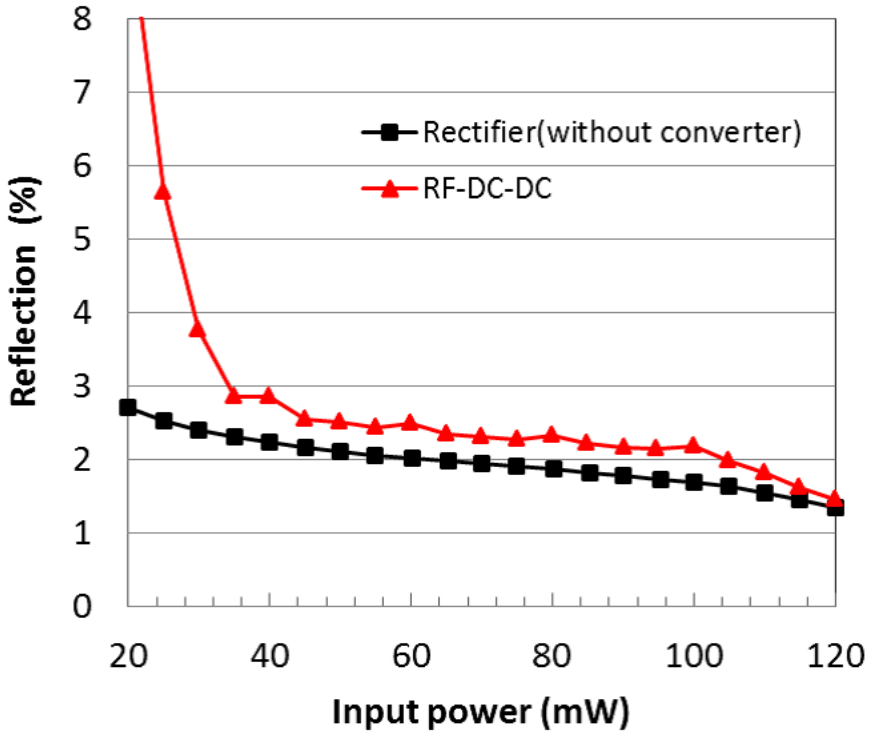


(b)

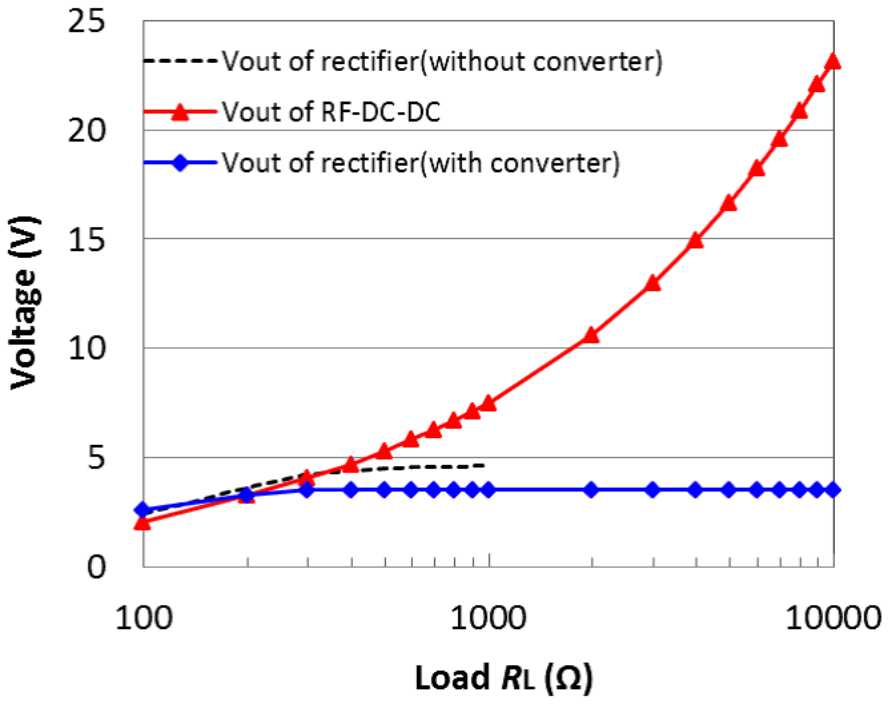
Figure 4.16: Reflection comparison between the general rectifier and the RF-DC-DC circuit. (a) Reflection vs. load resistance for a fixed input power of 80 mW. (b) Reflection vs. input power for a fixed load of 800 Ω.

Next, Figure 4.17 shows the output voltage of each single part of the RF-DC-DC circuit and rectifier (without converter). In Fig. 4.17(a), as expected, the V_{out} of the rectifier (with converter) is almost constant at 3.5 V because the variation of R'_{in} is sufficiently small. Observe that the V_{out} of the RF-DC-DC circuit, a boosted voltage, which well agrees with the curve in Fig. 4.11(a). In contrast, the V_{out} of the rectifier (without converter) is also given by the black point curve. In Fig. 4.17(b), the V_{out} of the rectifier (with converter) is increasing as the input power increases where the variation is as same as the one without converter, and the V_{out} of the RF-DC-DC circuit is increasing with the input power. Compared with the voltage conversion of the buck-boost converter (without rectifier), Figure 4.18 gives the output-input voltage curve of the buck-boost converter in the RF-DC-DC circuit. It is obvious that the two curves are in good agreement when the input voltage is higher than 2.5 V. The input voltage is boosted just a little by the converter because of a low efficiency at low input voltage level.

4.5 Summary

This chapter, firstly, a high-efficiency rectifying circuit used for 2.45 GHz microwave is designed and measured. Compared with the circuit designed in Chapter 3.4, this rectifier also uses a class-F load part while it reverses the polarity of the HF diode to obtain a positive output voltage and removes the DC block capacitor in the structure. The measured peak efficiency is 82.5 % at the optimal load of 190 Ω for an input microwave power of 80 mW.

Secondly, we design a self-powered non-inverting DCM buck-boost converter. Compared with the buck-boost converter designed in Chapter 3.3, there are mainly six improvements as follows. (1) It can obtain a positive output voltage with a positive input voltage just changing the position of MOSFET and diode. (2) An additional DC power source is needless because the control-pulse circuit is powered by the input voltage of the converter. (3) The input resistance of the buck-boost converter R'_{in} is designed to be equal to the optimal load of the rectifier (190 Ω). (4) The MOSFET gate driver is removed which attributes to a low power consumption of the control-pulse circuit. (5) A relatively big input capacitor (100 μF) is used to restrain the ripple of output voltage from the rectifier. (6) A variable resistance is used which makes the low frequency duty-on ratio be manual adjustable, then the input resistance R'_{in} is adjustable. Based on the simulation and experimental



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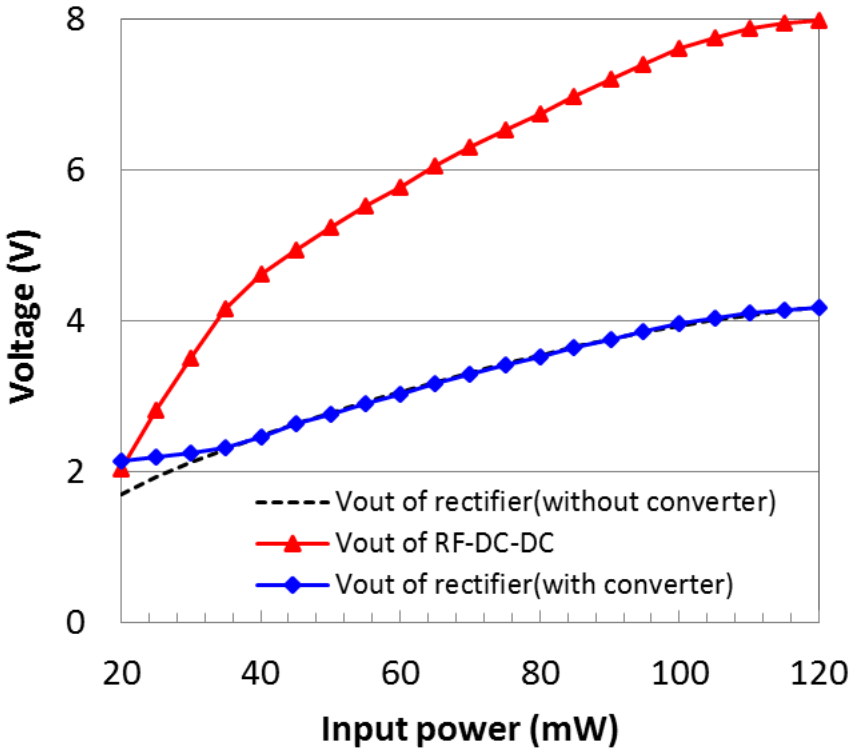


(b)

Figure 4.17: Output voltage of each part of the RF-DC-DC circuit. (a) Output voltage vs. load resistance for a fixed input power of 80 mW. (b) Output voltage vs. input power for a fixed load of 800 Ω.

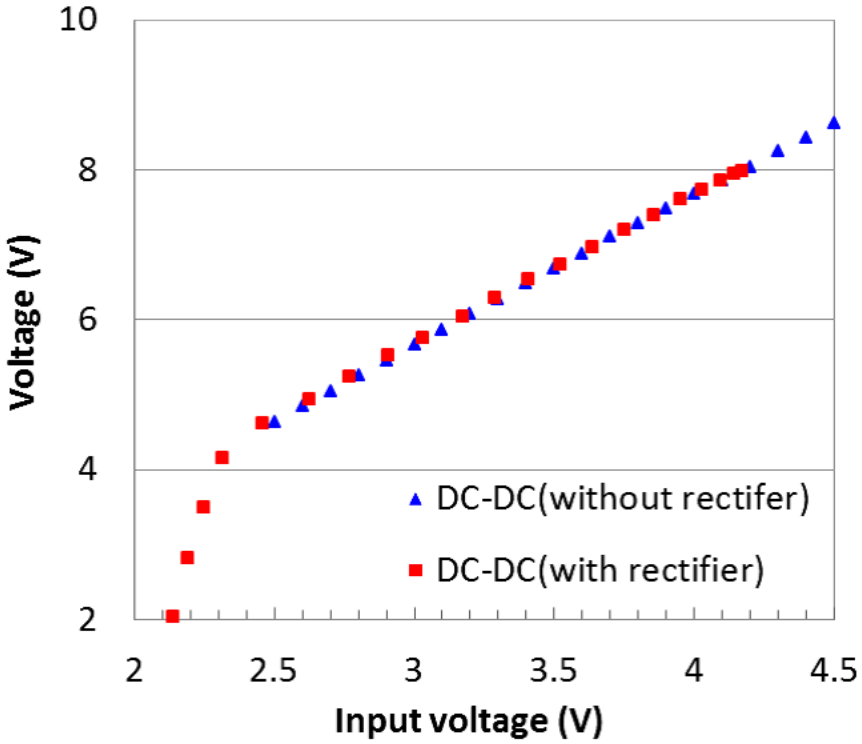


Figure 4.18: Input/output voltage relationships of the buck-boost converter in RF-DC-DC circuit.

results, the input resistance of the converter is almost constant at 190 Ω despite the changes of load resistances and input powers, which is in accordance with the theoretical analysis. Moreover, the measured efficiency of the converter exceeds 83 % for a wide load range of 200 Ω to 10 kΩ, and the input voltage is boosted by the buck-booster converter.

Finally, we conduct an experiment on RF-DC-DC circuit using the self-powered buck-boost converter and the rectifier. The conversion efficiency of the rectifier is approximately steady at 80 %, despite the load resistance varying from 200 Ω to 10 kΩ for an input microwave power of 80 mW. Compared with the peak efficiency of 82.5 % at the optimal load, the maximum power tracking efficiency is over 97 %. Furthermore, the overall efficiency of the RF-DC-DC circuit is constant over 66 % in an ultra-wide load range of 200 Ω to 10 kΩ (a ratio of 1:50). On the other hand, with a fixed load of 800 Ω, the overall efficiency of the RF-DC-DC circuit is also constant over 66 % for the input power of 40 to 120 mW. According to the experimental results, the proposed self-powered RF-DC-DC circuit exhibits a high efficiency in an ultra-wide load range.

A comparison between Figs. 3.14 and 4.13 shows that the varying ripple of

the overall efficiency is small in the self-powered RF-DC-DC circuit. Similarly, the varying ripple of the input resistance is also small in the self-powered RF-DC-DC circuit by comparison with Figs. 3.15 and 4.15. Make a comparison between externally-powered RF-DC-DC circuit and self-powered RF-DC-DC circuit in Table 4.4 where the superiors of the self-powered RF-DC-DC circuit are obvious.

In this Chapter, a self-powered RF-DC-DC circuit as an improved rectifying circuit, which exhibits a high efficiency in an ultra-wide load range, is developed. However, there are still some problems to be solved such as: (1) As the supplied voltage limitation from control-pulse circuit, the input RF power of the RF-DC-DC circuit is limited in the power range of 30 to 140 mW. It is expected to be used for a much more wide power range. (2) The efficiency of the buck-boost converter is expected to be much more high (such as 95 %) to reduce the loss caused by the converter, which would make the overall efficiency of the RF-DC-DC circuit close to the peak efficiency of the general rectifying circuit as much as possible. (3) The RF-DC-DC circuit is expected to be adapted to much small load resistance such as the range of 100 mΩ to 10 Ω. (4) This resistance conversion buck-boost converter is also expected to be valuable for other WPT system such as inductive coupling and magnetic or electromagnetic resonant coupling.

Table 4.4: Comparison between externally-powered and self-powered RF-DC-DC circuits. Z_{opt} is the optimal load of the rectifier. P_{RF} is the input RF power.

Item	externally-powered RF-DC-DC circuit	self-powered RF-DC-DC circuit
Power supply	require an extra DC power	supplied by the input power
$\eta_{\text{max(rf-dc)}}$	79.2 %	82.5 %
Z_{opt} and R_{in}	$R_{\text{in}} = 133\ \Omega$; $Z_{\text{opt}} = 160\ \Omega$.	$R_{\text{in}}^{'} \approx Z_{\text{opt}} = 190\ \Omega$
R_{in}	unadjustable	adjustable
V_{in} ripple	large	small
η_{track} vs. load	> 94 % in 200-5000 Ω (1:25)	> 97 % in 200-10000 Ω (1:50)
η_{track} vs. P_{RF}	> 93 % in 20-100 mW (1:5)	> 95 % in 20-100 mW (1:5)
η_{all} vs. load	> 60 % in 100-5000 Ω (1:50)	> 66 % in 200-10000 Ω (1:50)
η_{all} vs. P_{RF}	> 61 % in 40-100 mW (1:2.5)	> 66 % in 40-120 mW (1:3)

Chapter 5

Experiment on Driving a DC Motor Using MPT

5.1 Introduction

In this chapter, the RF-DC-DC circuit proposed previous is used to drive a dynamic load resistance device. We conduct several experiments on driving a low-power DC motor using microwave power transmission. As the self-powered buck-boost converter in Chapter 4 exhibits a low efficiency at low input power level, we design a relatively high output voltage rectifier and an improved buck-boost converter whose control-pulse circuit is supplied by the output voltage of the converter. By comparison, we also design a relatively low output voltage rectifier and an improved buck-boost converter whose control-pulse circuit is supplied by the input voltage of the converter. This chapter proposes a compact power-receiving device which consists of a rectenna array and the improved self-powered buck-boost converter. With the combination of two different rectenna arrays and two different converters, there are four different power-receiving devices. Finally, using those power-receiving devices, some MPT experiments on driving the DC motor using continuous-wave (CW) and pulsed-wave are conducted.

5.2 Background

The Mars Exploration Rover mission is part of NASA's Mars Exploration Program, a long-term effort of robotic exploration of the red planet. Primary among the mission's scientific goals is to search for and characterize a wide range of rocks and soils that hold clues to past water activity on Mars. The spacecraft are targeted to sites on opposite sides of Mars that appear to have been affected by liquid water in the past. NASA's twin robot geologists, the Mars Exploration Rovers, launched toward Mars on June 10 and July 7, 2003, in search of answers about the history of water on Mars. They landed on Mars January 3 and January 24 PST, 2004 [74]. Power supply technology for rovers is an important issue. So far, several means for electric power supply are used for Mars exploration rover, such as solar cells, lithium batteries and a radioisotope power system. However, there are some problems such as the increase of weight and the limitation of activity time and area. For example, using solar power limits the places on Mars that the rover can explore. They are restricted to landing and traveling around the equatorial region where they can get enough sunlight to re-energize their batteries. Therefore, MPT, an innovative way for powering moon or planet exploration rovers, is proposed [75]. Similarly, for a wide area and continuous Mars observation, MPT for Mars exploration airplane is also proposed [26].

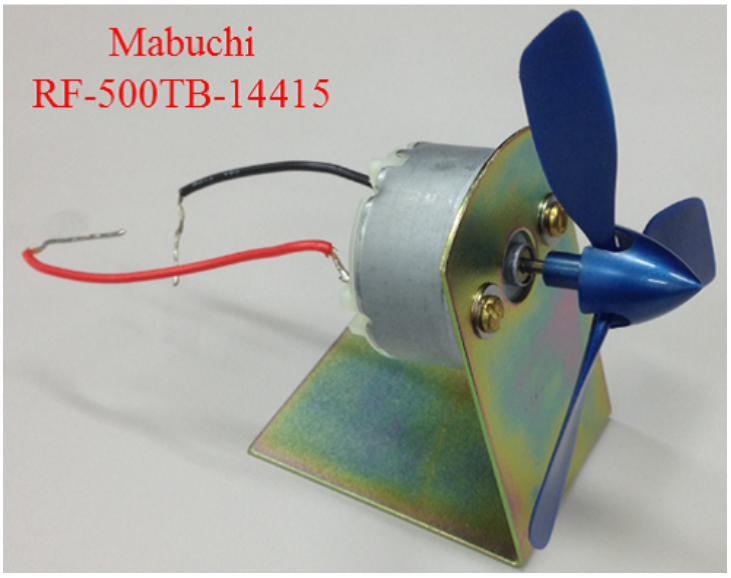
Both rover and airplane proposed above are mainly actuated by DC motors. A static solar power station generates DC power and changes it into microwave power, then transmits microwave power to a distant rover or airplane. The rover or airplane, equipped with rectenna (or rectenna array), receives microwave power and changes it into DC power, which actuates DC motors of rover or a propeller of airplane. As we know, input V-I characteristic of DC motor exhibits variable which depends on its revolution speed and mechanical load. Therefore, the emulated input resistance of DC motor is variable. As a result, the rectifying efficiency is variable when the rectenna is connected to a DC motor. The motor would not work well even stop when it is changed from one operating condition to another because of power shortage caused by impedance mismatch.

Therefore, to solve the problems mentioned above, this chapter only focuses on the study of driving a DC motor using MPT. We design a power-receiving device which has capability of efficiently driving a dynamic load resistance device (such as a DC motor) in the MPT system.

5.3 Low-power DC Motor

If the load resistance characteristics of a DC motor match the optimal load of the rectenna, the output characteristics of the rectenna will be equivalent to those when the rectenna is loaded by a load resistance [38]. It means that a rectenna can drive a DC motor with the maximum efficiency when the load characteristic of the motor equals to the optimal load of the rectenna. However, powered by the microwave, a motor is difficult to maintain a steady revolution speed due to the change of mechanical loads, which leads to a decrease in efficiency of the rectenna. Therefore, if we insert a constant input resistance buck-boost converter between the rectenna and the DC motor, the output impedance of the rectenna maintains constant despite operating conditions of the motor. Consequently, the efficiency of the rectenna will keep constant. Hence, it is necessary to check the input V-I characteristics or the input resistance characteristics of a DC motor before designing a rectenna.

The DC motor used in this study is an RF-500TB-14415 motor made by Mabuchi Motor Co., Ltd as shown in Fig. 5.1(a). This type of motor is usually used for low-power household electrical appliances such as audio-video equipment. The operating voltage range is from 1.5 to 9 V and the output power is widely from 0.01 to 2 W [76]. A DC power source is used to measure the input load characteristics of the motor connected with a small fan as a reference mechanical load. The block diagram of the measurement circuit is shown in Fig. 5.1(b), where an ammeter and a voltmeter measure the input current I_{m} and the input voltage V_{m}, respectively. Then the emulated input resistance R_{m} and input power P_{m} can be calculated by $R_{\mathrm{m}} = V_{\mathrm{m}}/I_{\mathrm{m}}$ and $P_{\mathrm{m}} = V_{\mathrm{m}}I_{\mathrm{m}}$, respectively. To determine R_{m} and P_{m}, the input current is measured with a changed input voltage. The results are shown in Fig. 5.2, which indicates that the input resistance R_{m} increases from 36 to 140 Ω and the input power P_{m} increases from 10 to 645 mW when the input voltage increases from 0.6 to 9 V. Of course, instead of the small fan, if a heavy mechanical load was used, the input resistance would be lower because of a higher current consumption. As the same reason, R_{m} decreases at a higher input voltage. When the input voltage exceeds 5 V, the revolution speed of the fan becomes very fast which results in a heavy resistance of air, then the increase of the current consumption leads to the decrease of R_{m}.



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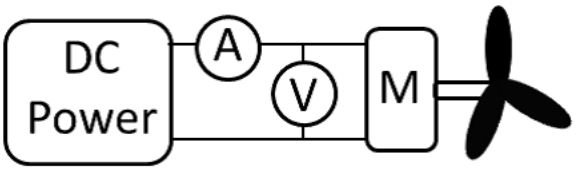


(b)

Figure 5.1: Photograph of the DC motor and measurement setup for input load resistance R_{m}. (a) Photograph. (b) Measurement setup. An ammeter and a voltmeter measure the input current I_{m} and the input voltage V_{m}, respectively.

5.4 Rectenna Design

5.4.1 Antenna Design

The rectenna is very important for converting wireless RF power into DC power. A rectenna usually contains a receiving antenna and a rectifying circuit. The receiving antenna collects microwave incident power and the rectifying circuit converts microwave power into DC power. In the previous chapters, we only focused on the discussion about rectifying circuits. This chapter will discuss about the design of an antenna.

An antenna is an electrical device for radiating or receiving radio waves. On transmission, an antenna accepts energy from a transmission line and radiates it into space, and on reception, an antenna gathers energy from an incident wave

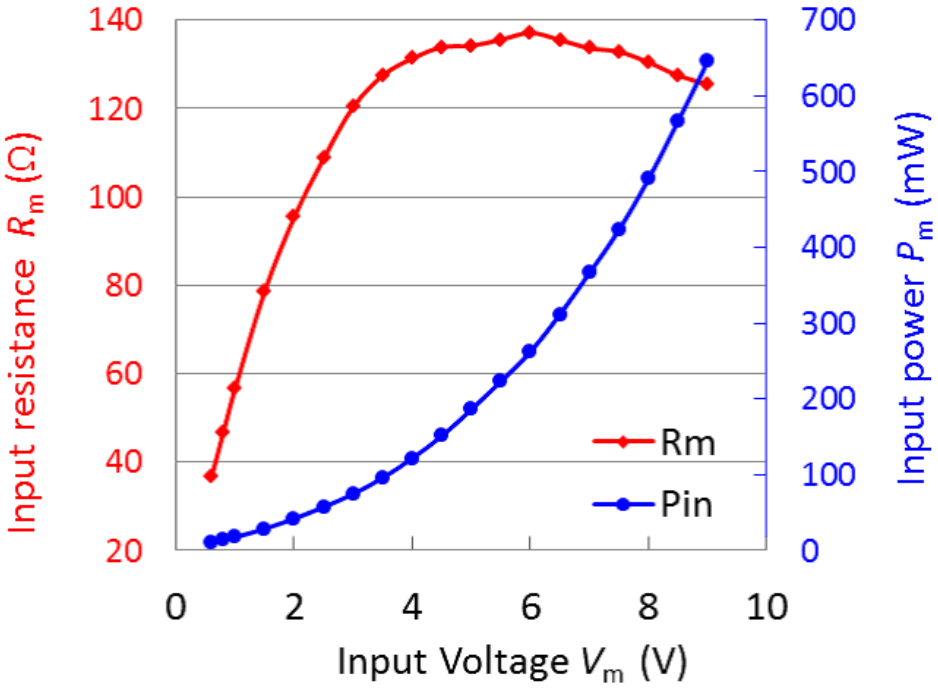


Figure 5.2: Measured input load characteristics of the DC motor.

and sends it down a transmission line. In this chapter, an antenna designed to receive microwave from transmission side. According to different shape and application, there are linear antenna, planar antenna, aperture antenna and so on. For microwave power transmission, a microstrip patch antenna (or array) is commonly used by many researchers [13]-[15]. Patch antenna, with a simple shape, performs a relatively high gain so that it can receive much more power in a limited microwave power density environment. In addition, microstrip patch antennas are excellent in low fabrication cost and readily amenable to mass production [77].

In this experiment, since a horn antenna is used as a microwave transmission antenna, we design a linear polarization circular patch antenna as a receiving antenna. With respect to the dielectric substrate for the rectifier circuit, we all use the same dielectric substrate of NPC-F260 as shown in Table 3.4, where the thickness of the substrate H is 0.8 mm in previous chapters. However, the thickness of the substrate decreases resulting in dielectric losses increasing so that the gain of the patch antenna will decrease, which has been reported in the literature [78][79]. In this experiment, it requires a high gain for receiving as much power as possible in a limited power density environment. Therefore, we design two antennas with two different thickness of dielectric substrates (NPC-F260). Here, these two antennas such as H = 1.6 mm and H = 0.8 mm are referred to as H16 and H08, respectively.

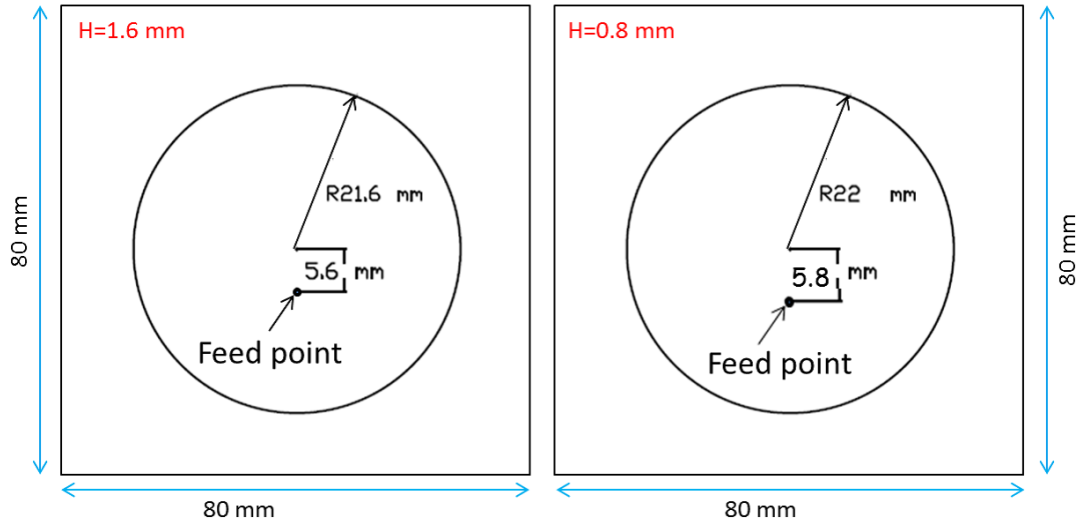


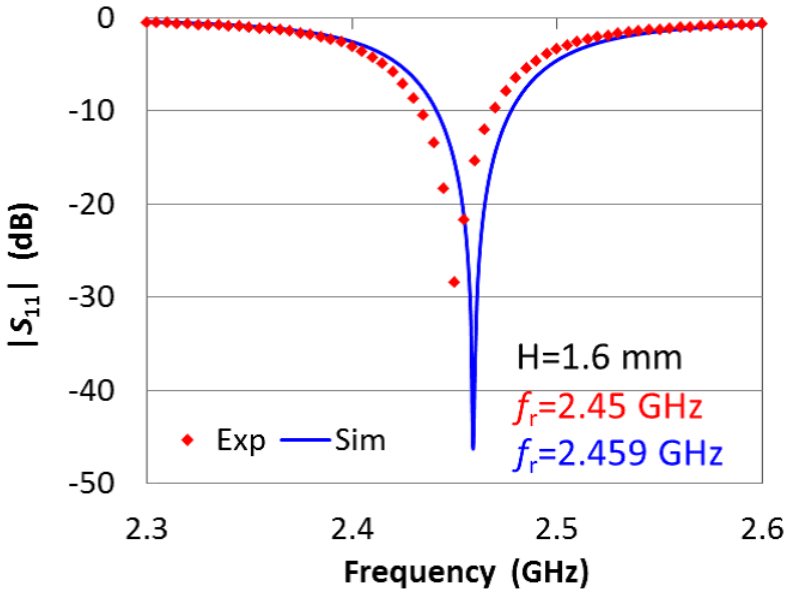
Figure 5.3: Designed antenna sizes.

We simulate two antennas by using HFSS simulator of ANSYS and fabricate them. Figure 5.3 shows the shape and sizes of antennas, where the whole size are both 80 mm × 80 mm, and the radius are 21.6, 22 mm, the distances from center to feed point are 5.6, 5.8 mm, respectively. Figure 5.4 shows the simulation and measured results of $|S_{11}|$. Note that the simulation results of two antennas are both a little different with the measured results. However, we make sure that the measured results are both resonant at the frequency of 2.45 GHz, where the $|S_{11}|$ of H16 and H08 are -28 dB and -24 dB, respectively.

The gain of an antenna is a very important factor in rectenna. The simulation results of antenna's gain are shown in Fig. 5.5, where the gain of H16 and H08 antennas are 7.04 dBi and 6.24 dBi, respectively. Then, as for measurement, compared to a standard half-wave dipole antenna with a known gain of 2 dBi, the gain of designed antenna can be calculated. We set the same power density condition to measure the receiving power of two designed antennas and the standard dipole antenna, then the gain of each antenna can be calculated referring to the receiving power. The effective aperture of the antenna A_e can be calculated by following:

$$A_e = \frac{G_A \lambda^2}{4\pi}, \tag{5.1}$$

where, G_A is the power gain, and λ is the free-space wavelength at 2.45 GHz. The results can be seen from Table 5.1. Note that the effective aperture of the H16 antenna is bigger than that of the H08 antenna. This means one H16 antenna can



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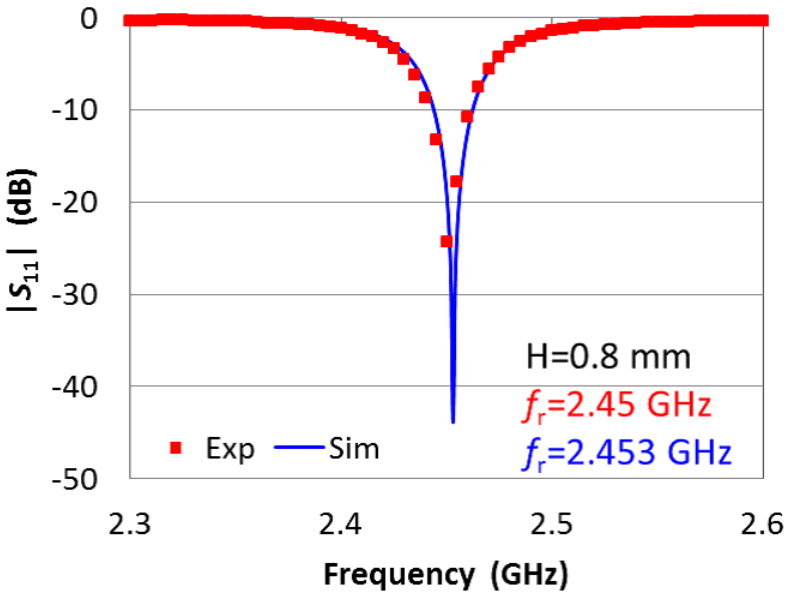


(b)

Figure 5.4: Simulated and measured $|S_{11}|$ of antennas. (a) Antenna H16 $|S_{11}|$ = -28 dB. (b) Antenna H08 $|S_{11}|$ = -24 dB.

Table 5.1: Comparison of H16, H08 and standard antennas for receiving power, gain and effective aperture.

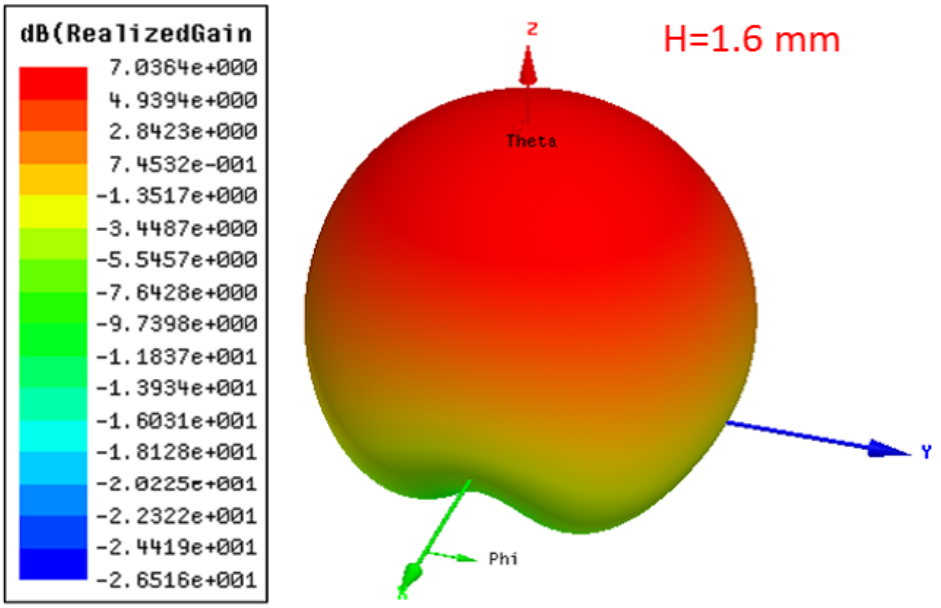
Antenna	H16	H08	Standard
Receiving power (mW)	13.53	11.1	4.92
Gain (dBi)	6.39	5.53	2
Effective aperture (cm^2)	51.6	42.3	18.8

receive approximately 52 mW while H08 antenna can receive approximately 42 mW at the same microwave density of 1 mW/cm^2 which is the standard value of the radiation protection guidance. Consequently, we choose the H16 antenna as a receiving antenna because it is superior in capability of receiving power.

5.4.2 Small Rectifying Circuit

In order to make a compact rectenna, we design a small rectifying circuit to match with the size of antennas. As mentioned in Chapter 4.5, the output voltage of the rectifying circuit supplies voltage for the control-pulse circuit of the buck-boost converter. The overall efficiency of the RF-DC-DC circuit is low when the input power is below 40 mW because the output voltage of the rectifier is low so that it results in a low efficiency of the buck-boost converter. Therefore, it demands a higher output voltage in a lower input power level. On the other hand, a higher RF-DC conversion efficiency is also required. Considering all requirements mentioned above, we design a small rectifier A with a higher optimal load resistance to get a higher output voltage at the same power level. Compared with rectifier A, a small rectifier B with a normal optimal load resistance is also designed.

We use the same dielectric substrate of NPC-F260 and the thickness is 0.8 mm which is the same as the previous design. Similarly, an input matching circuit and two HSMS286F HF diodes are used in the rectifier, while the output filter consists of several open stubs without $\lambda/4$ line. Here, with respect to 2.45 GHz microwave in the NPC-F260 substrate, $\lambda/4$ line is approximate 21 mm which is almost the half length of the whole circuit as shown in Fig. 3.11. $\lambda/4$ line mainly addresses the fundamental frequency content of the output voltage. If we sufficiently deal with the other high order harmonics, the overall high frequency



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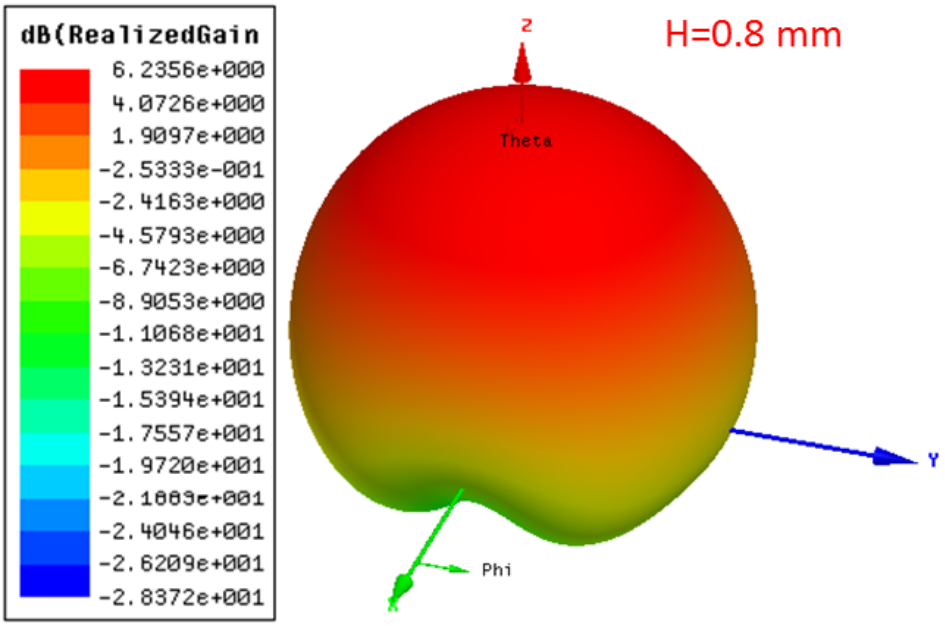
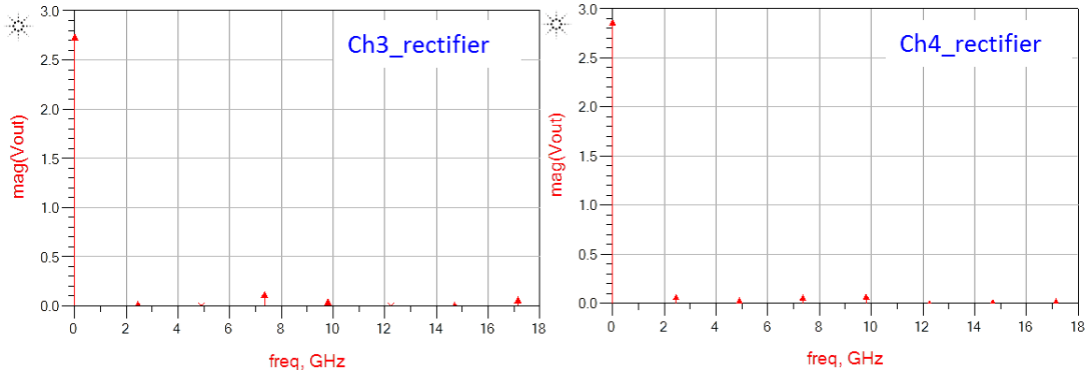


(b)

Figure 5.5: Simulated gains of antennas. (a) The gain of H16 antenna is 7.04 dBi. (b) The gain of H08 antenna is 6.24 dBi.

harmonic loss also can be restrained at a low level even without $\lambda/4$ line. Furthermore, the losses of high frequency harmonic contents are small for 2.45 GHz microwave. Figure 5.6 shows the simulation results of output voltage spectrum of four rectifiers. In Fig. 5.6(a), Ch3_rectifier and Ch4_rectifier are the rectifiers designed in Chapter 3 and Chapter 4, respectively. Figure 5.6(b) shows the results of the rectifier A and B. Observe that the fundamental frequency content in A and B are bigger than that in Ch3_rectifier and Ch4_rectifier, while the high order harmonics of the former is smaller than the latter one. As a result, the overall high frequency harmonic loss of the former is approximately equivalent to the latter one.



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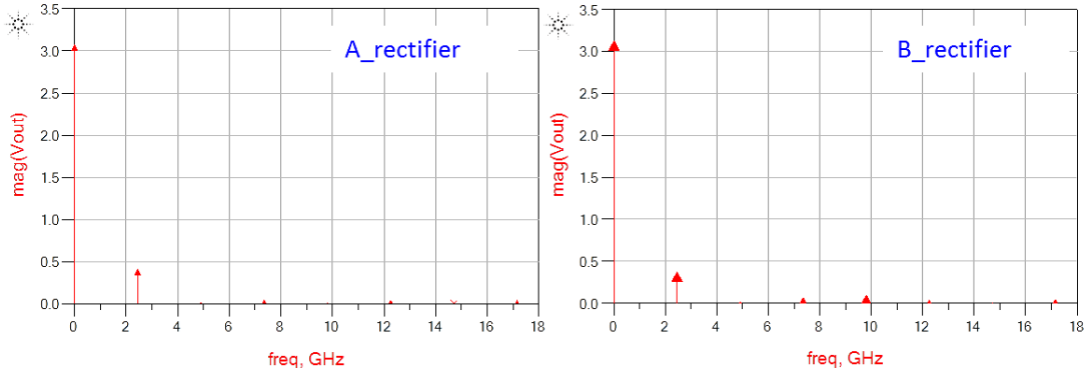


(b)

Figure 5.6: Simulated output voltage spectrum of rectifiers. (a) Ch3_rectifier and Ch4_rectifier are the rectifiers (with $\lambda/4$ line) designed in Chapter 3 and Chapter 4, respectively. (b) A_rectifier and B_rectifier are the rectifiers (without $\lambda/4$ line) designed in this chapter.

Figure 5.7 shows the photograph of the designed rectifier A, B and Ch4_rectifier as a contrast. The size of both rectifier A and B are 35 mm×50 mm, which is

almost the half size of the Ch4_rectifier. Figure 5.8 shows the measured efficiency-load characteristics of the rectifier A and B at different input power levels. The results clearly indicate that the efficiency of the rectifier depends on the connected load and the input power. As for rectifier A, the optimal load is almost constant at 300 Ω in the input power range of 10 to 60 mW. The maximum efficiency is 80.8 % for a load of 300 Ω and an input power of 60 mW. As for rectifier B, the optimal load is almost constant at 160 Ω in the input power range of 10 to 60 mW. The maximum efficiency is 80 % for a load of 160 Ω and an input power of 80 mW. It is clear that the optimal load of the rectifier A is almost twice as big as that of the rectifier B.

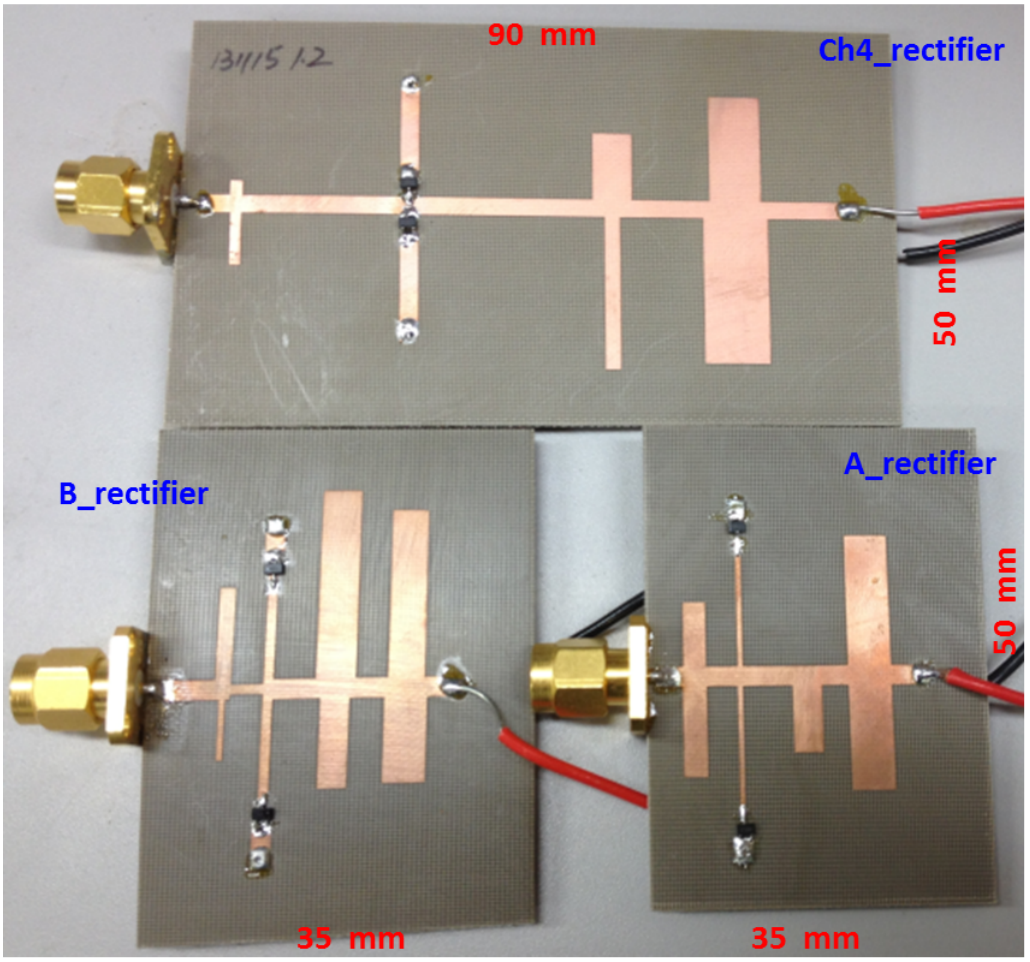
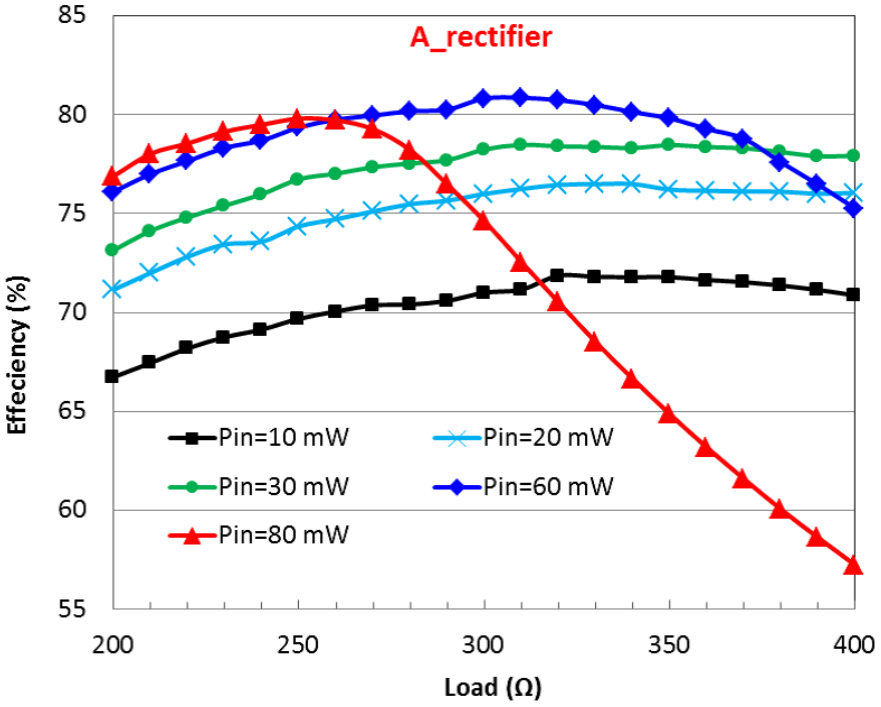


Figure 5.7: Photograph of designed rectifiers: A, B and Ch4_rectifier.

We also investigate the output voltage of both rectifiers and compare them in Fig. 5.9, where A_Eff, A_Vo are the efficiency and output voltage of the rectifier A, B_Eff, B_Vo are the efficiency and output voltage of the rectifier B. Figure 5.9(a) compares the efficiency and output voltage versus load resistance at a fixed input



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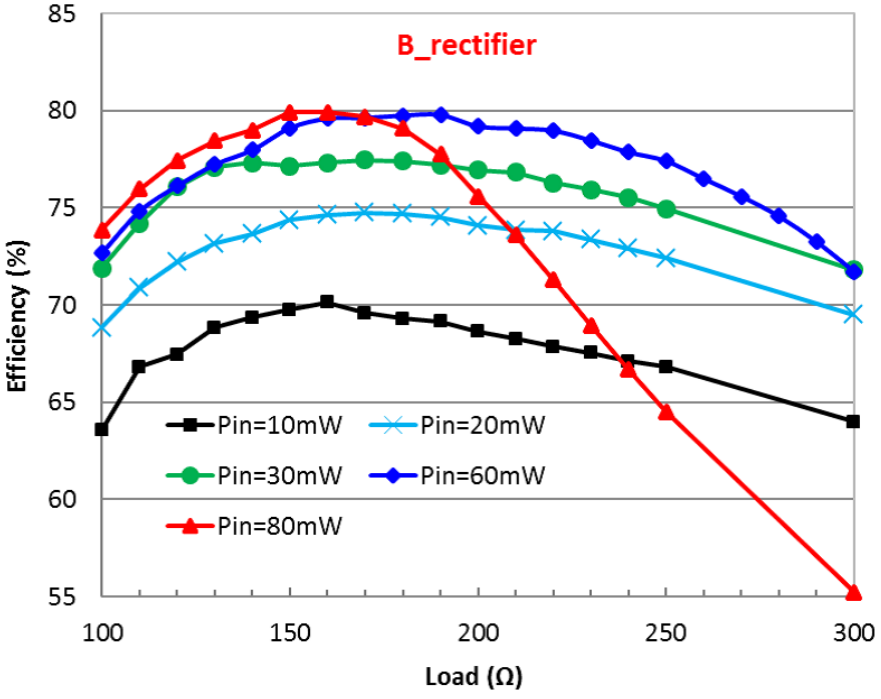


(b)

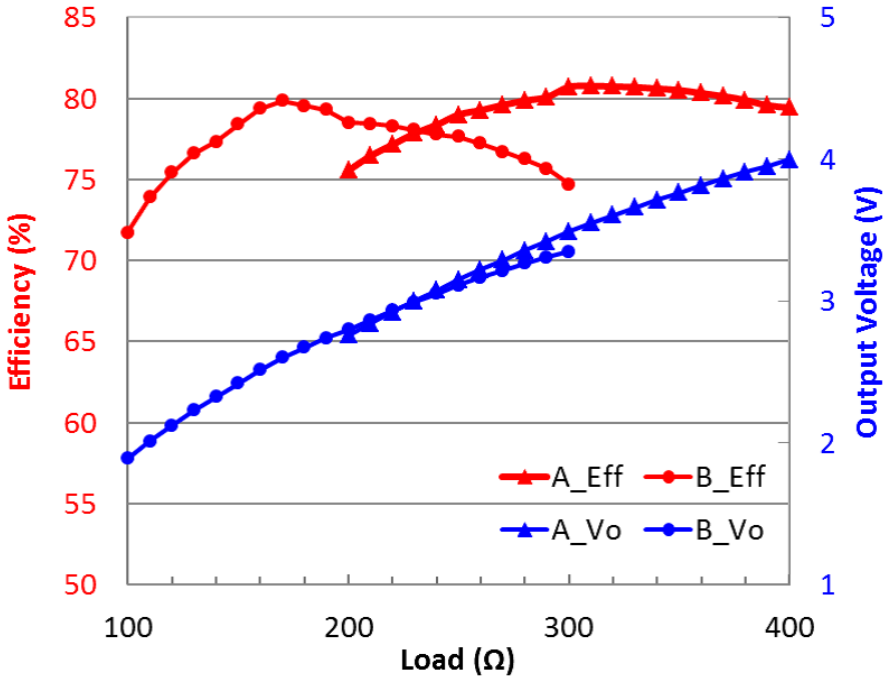
Figure 5.8: Measured efficiency results of the rectifier A and B for different input powers. (a) Efficiency-load characteristics of the A rectifier. The optimal load is almost 300 Ω. (b) Efficiency-load characteristics of the B rectifier. The optimal load is almost 160 Ω.

Table 5.2: Comparison of four rectifiers. Z_{opt} presents the optimal load of the rectifier. V_{out} (80 mW and Z_{opt}) means the output voltage of the rectifier at the optimal load for the same input power of 80 mW. −3.2 V means that the Ch3_rectifier obtains a negative output voltage which has been mentioned in Chapter 3.

Rectifier	Ch3	Ch4	A	B
Peak efficiency	79.2 %	82.5 %	80.8 %	80 %
Z_{opt} (Ω)	160	190	300	160
V_{out} (80 mW and Z_{opt}) (V)	−3.2	3.5	4.3	3.2
Size (mm)	90×50	90×50	35×50	35×50

power of 50 mW. Observe that the rectifier A presents a higher efficiency and output voltage at a higher load resistance while the rectifier B exhibits a higher efficiency and output voltage at a lower load resistance. In Fig. 5.9(b), when the input power is changing from 10 to 80 mW, the output voltage of both rectifiers are increasing while rectifier A obtains a higher output voltage than rectifier B, the efficiency of the rectifier B is increasing and levels off, while the efficiency of the rectifier A is increasing at the beginning and decreases at a higher power level. The reason is as following. Rectifier A obtains a higher output voltage which leads to a higher reverse voltage applied on the rectifying diode. Consequently, the reverse voltage applied on the diode exceeds the breakdown voltage at a lower input power level which results in efficiency decreasing of the rectifier A.

Until now, we have designed and measured four rectifiers which have each merits and demerits as summarized in Table 5.2. Note that the higher optimal load rectifier obtains a higher output voltage and the output voltage of the A rectifier is 1 V bigger than that of the B rectifier. Ch3_rectifier and rectifier B have almost the same peak efficiency and the same optimal load while the size of the rectifier B is half smaller than Ch3_rectifier. Compared to Ch4_rectifier, the efficiency of the rectifier A is slightly lower but it obtains a higher output voltage with a half small size.



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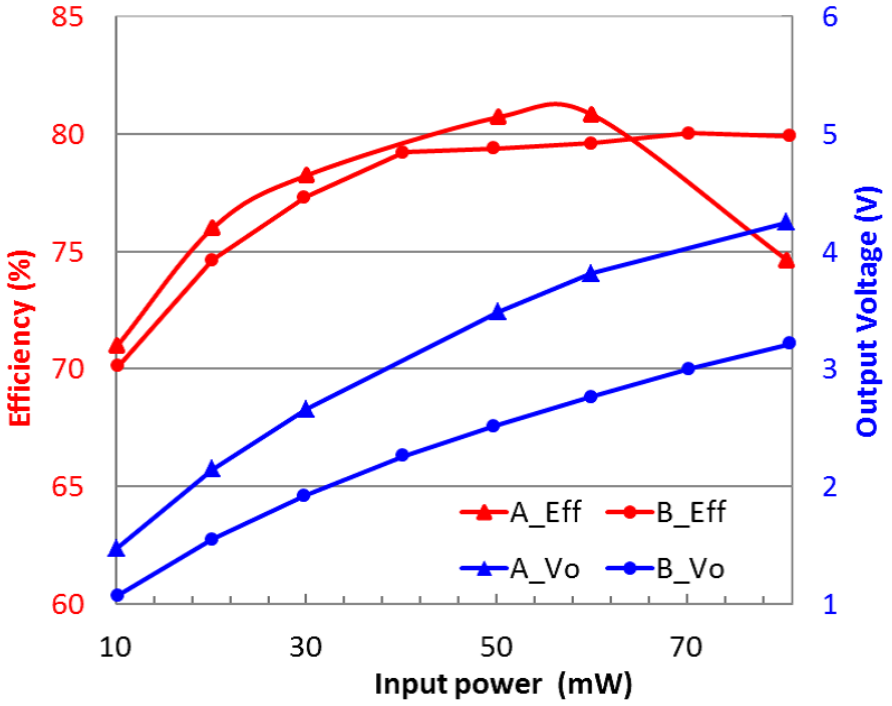


(b)

Figure 5.9: Comparison of the measured efficiency and the output voltage. (a) Efficiency and output voltage vs. load for a fixed input power of 50 mW. (b) Efficiency and output voltage vs. input power for the optimal load of 300 Ω (rectifier A) and 160 Ω (rectifier B), respectively. A_Eff, A_Vo are the efficiency and output voltage of the rectifier A; B_Eff, B_Vo are the efficiency and output voltage of the rectifier B.

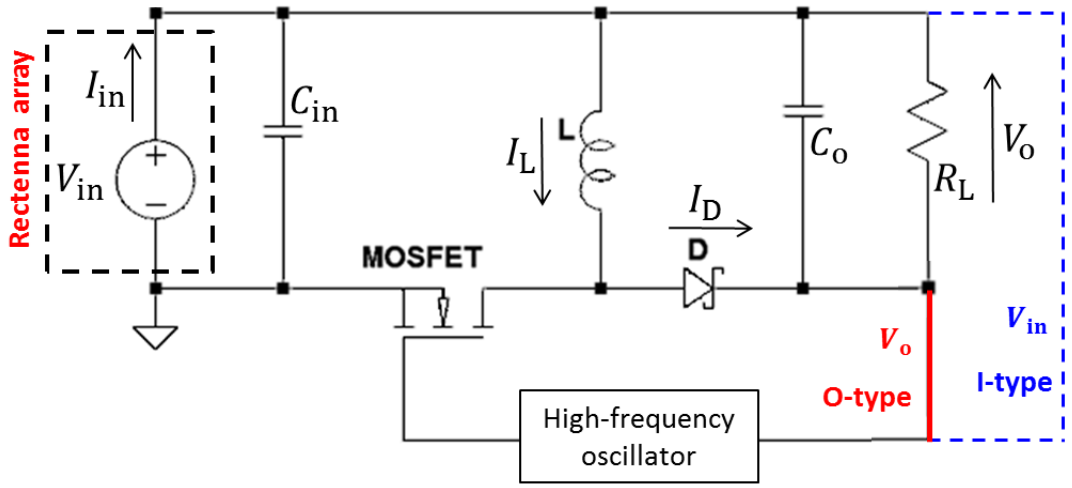
5.5 Compact Power-receiving Device

5.5.1 Improved Buck-boost Converter

A buck-boost converter with a constant input resistance is previously proposed to improve the efficiency-load characteristics of the rectifier in Chapter 4.3. However, the converter operates at low efficiency for input power levels below 40 mW, because the control-pulse oscillator can not be powered by the input voltage of the converter at such low power levels. To overcome this issue, an improved buck-boost converter is designed as shown in Fig. 5.10. Compared to the previous one in Chapter 4.3, the improved buck-boost converter is smaller and simpler because it uses only HF oscillator and a small inductor L (220 μH). As seen the operating waveforms from Fig. 5.10(b), the operating condition has no sleep-state. More importantly, the power for the control-pulse oscillator is supplied by the output voltage of the converter, which is referred to as O-type. By comparison, the control-pulse oscillator is powered by the input voltage of the converter is also designed, which is referred to as I-type. As we have mentioned, the buck-boost converter designed in Chapter 4.3 exhibits a boost converter where the output voltage is larger than input voltage. Therefore, O-type converter is suitable for lower power level because of a higher supplied voltage for the control-pulse oscillator.

Similarly, the input resistance of the converter R_{in} can be calculated using Eq. (3.37). The other parameters are as follows: $L = 220$ μH, f_{h}=20 kHz, D_{h}=0.5, D_{ι} should equal 1 because there is no low-frequency oscillator in this case. This gives an R_{in} value of 35 Ω. According to Eq. (3.37), the input resistance R_{in} of this buck-boost converter will remain a constant value of 35 Ω even though the connected load resistance is above 35 Ω. Hence, when the DC motor described in Chapter 5.3 is connected to the buck-boost converter as a load, its input load resistance of 36 to 140 Ω satisfies the load requirements of the buck-boost converter. The components of the circuit used in this converter are given in Table 5.3 and the photograph of fabricated circuits are shown in Fig. 5.11, which has a size of 25 mm$\times$25 mm. For both converters, the components of the circuit used are the same and the only difference is the connection method.

Similarly, we use the measurement circuit in Fig. 4.9(a) to evaluate the efficiency and R_{in} of the buck-boost converter with a DC power source, and the



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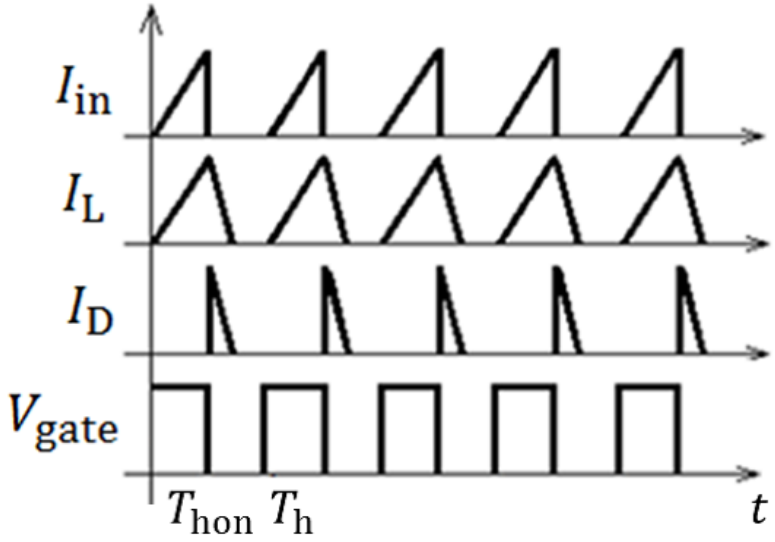
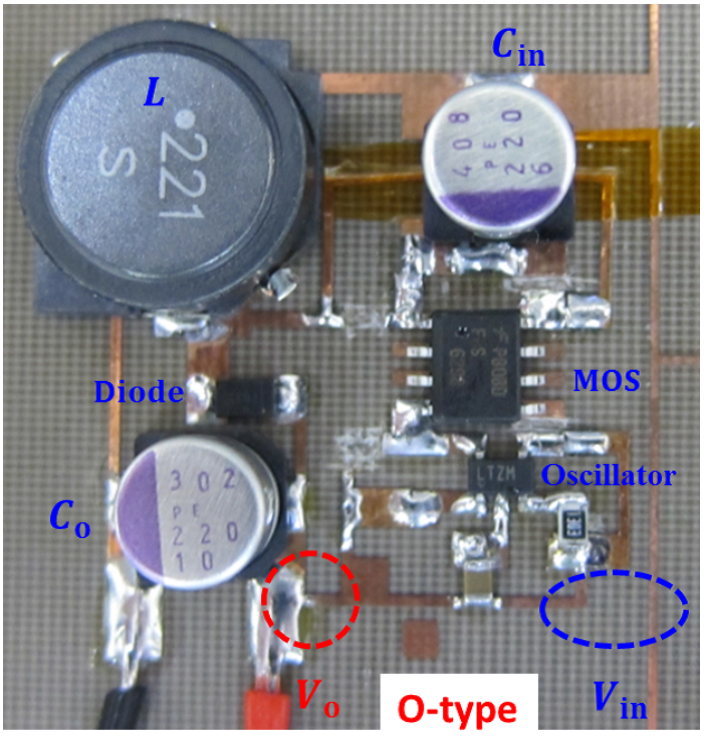


(b)

Figure 5.10: Block diagram and operating waveforms of the improved buck-boost converter. (a) Block diagram. (b) Operating waveforms. O-type: oscillator is powered by output voltage; I-type: oscillator is powered by input voltage.



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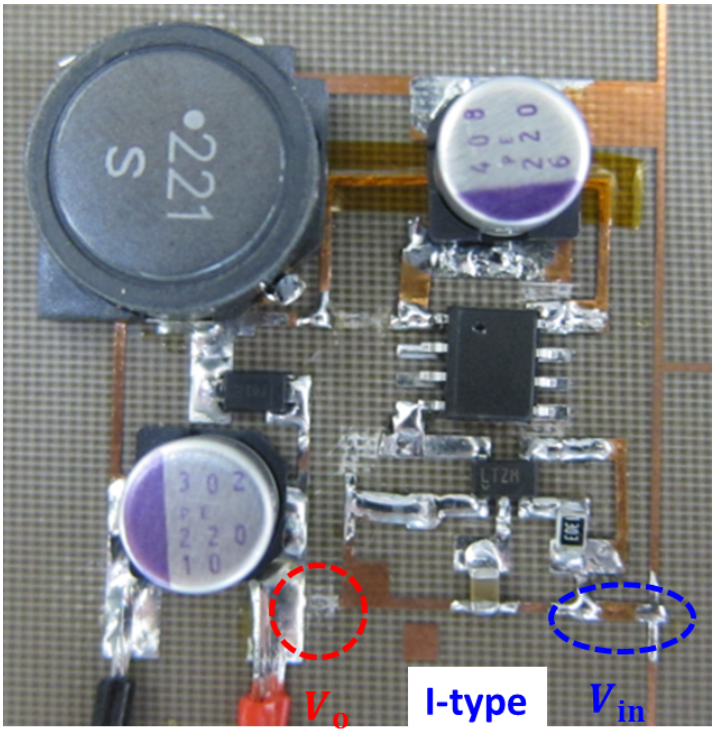


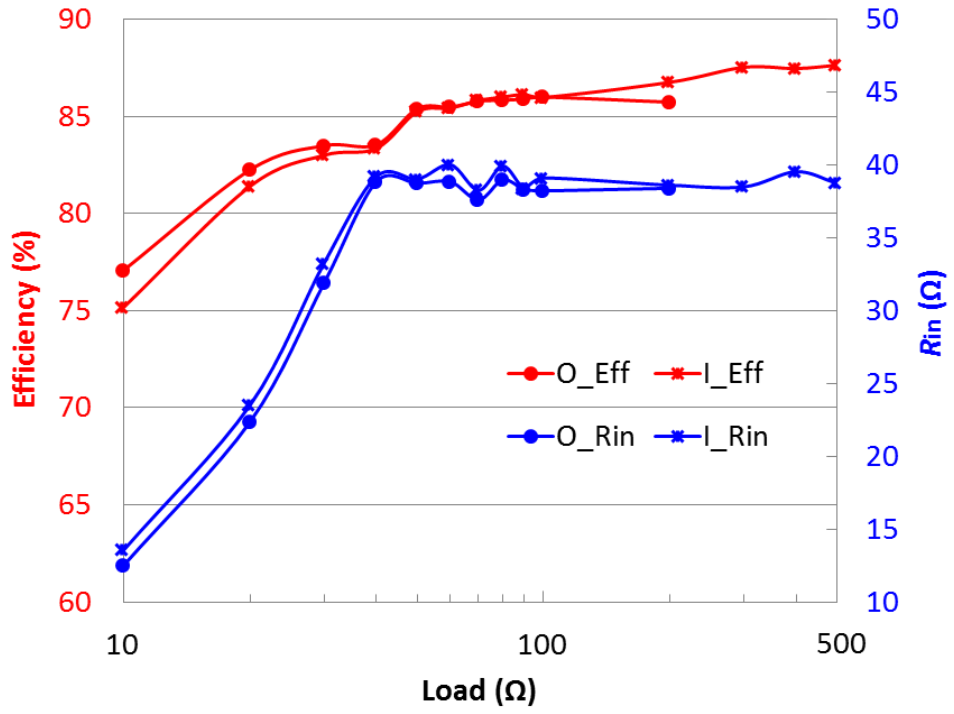
(b)

Figure 5.11: Photograph of fabricated buck-boost converters. (a) O-type: V_o line is connected to the oscillator and V_{in} line is open. (b) I-type: V_o line is open and V_{in} line is connected to the oscillator.

Table 5.3: Components of the improved buck-boost converters.

Component	Maker	Part NO.	Description
High-frequency oscillator	Linear Technology	LTC6900	$I_{SS} = 0.5$ mA ($f = 3$ MHz); $V_{SSmin} = 2.7$ V
N-MOS	Fairchild	FDS6298	$V_{DSS} = 30$ V; $R_{DS(on)} = 12$ mΩ ($V_{GS} = 4.5$ V)
Schottky Diode	Diodes Incorporated	DFLS130L	$V_{RRM} = 30$ V; V_{FM}= 0.21 V
Inductor	Taiyo Yuden	NS12555T221	$L = 220$ μH; $R_{dc} = 0.27$ Ω; $I_{dc} = 1$ A
Capacitor C_{in}	Panasonic	6SVPE220M	220 μF ($V_{rated} = 6.3$ V)
Capacitor C_o	Panasonic	10SVPE220M	220 μF ($V_{rated} = 10$ V)

comparison results of two type converters are shown in Fig. 5.12. Here, O_Eff, I_Eff are the efficiency of O-type and I-type converters, respectively. O_Rin, I_Rin are the input resistance of O-type and I-type converters, respectively. In Fig. 5.12(a), at a fixed input voltage of 3 V, the efficiency and R_{in} of the I-type are measured with load changing from 10 to 500 Ω, while those of the O-type are measured with load changing from 10 to 200 Ω in order to avoid the oscillator of the O-type over-voltage. The efficiency and R_{in} variation tendencies are approximately the same. The efficiency of the O-type is slightly better than the one of the I-type when the load is below 50 Ω and that relation is inverted when the load is over 100 Ω because the supplied voltage for oscillator is simultaneously varying with the output voltage of the O-type converter. As expected, the R_{in} is constant to be 38 Ω which is close to the designed value of 35 Ω when the load is over 40 Ω. In Fig. 5.12(b), at a fixed load of 50 Ω, the efficiency and R_{in} of the I-type are measured with input voltage changing from 2.1 to 5 V, while that of O-type is measured with input voltage changing from 1.1 to 3 V. It is obvious that the O-type converter has the capability of handing lower input voltage and the I-type converter has the capability of handing higher input voltage. Additionally, the R_{in} of both type converters are constant and independent of the input voltage.



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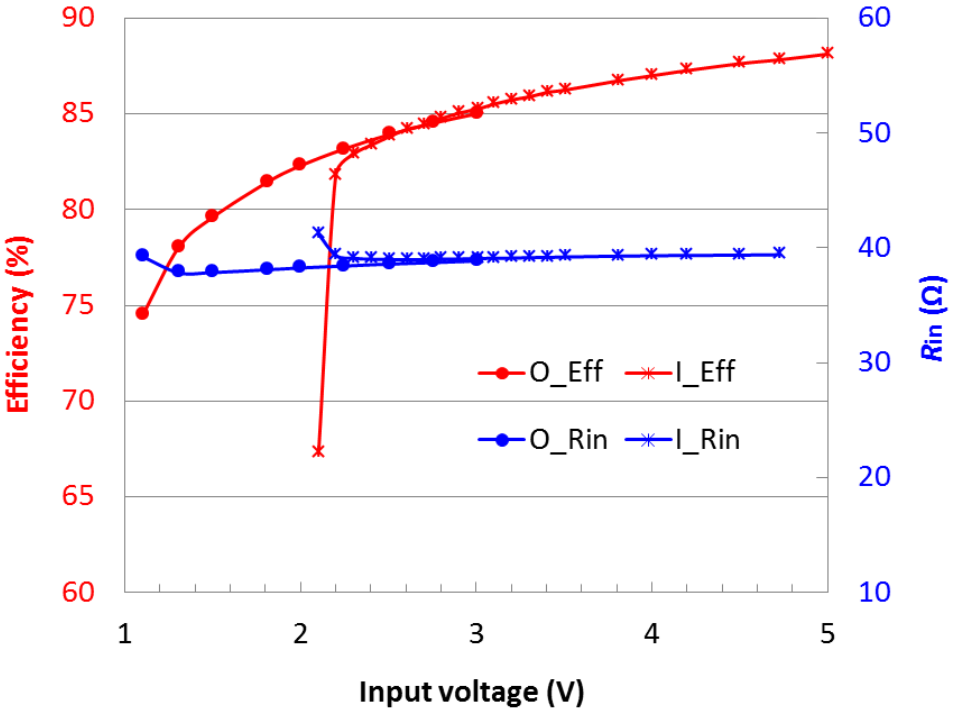


(b)

Figure 5.12: Comparison of measured results between O-type and I-type buck-boost converters. (a) Efficiency and R_{in} vs. load for a fixed input voltage of 3 V. (b) Efficiency and R_{in} vs. input voltage for a fixed load of 50 Ω.

Table 5.4: Comparison of four buck-boost converters such as the converter in Chapter 3, Chapter 4 and O-type, I-type converters in this chapter.

Converter	Ch3	Ch4	I-type	O-type
Power supply	extra DC power source	input voltage	input voltage	output voltage
Voltage polarity	inverting	non-inverting	non-inverting	non-inverting
Control circuit	HF, LF oscillators and gate driver	HF, LF oscillators	HF oscillator	HF oscillator
Size	large	medium	small	small

Until now, we have designed and measured four buck-boost converters which have each merits and demerits as summarized in Table 5.4. Those four converters all operate at discontinuous conduction mode with constant input resistance characteristic and play a boost converter role in the application. The O-type and I-type have the small size and O-type has the capability of handing lower power without extra DC power supply.

5.5.2 Compacting Rectenna Array and Buck-boost Converter

A single rectenna consists of the antenna designed in Chapter 5.4.1 and the rectifier as shown in Fig. 5.13. The optimal load of the A rectifier and B rectifier described in Chapter 5.4.2 are 300 Ω and 160 Ω, respectively. While the input resistance of the buck-boost converter described in Chapter 5.5.1 is almost 35 Ω. Therefore, in order to match the input resistance of the buck-boost converter, a common method that several rectennas are connected in parallel as a rectenna array is usually applied to decrease the optimal load of the rectenna [80]. Then we set 9 A_rectennas connected in parallel as A rectenna array and 4 B_rectennas connected in parallel as B rectenna array, respectively. The optimal load of the A rectenna array and the B rectenna array can be calculated to be approximately 33 Ω and 40 Ω, respectively.

Additionally, in order to make the rectenna array in a compact board, we use a multilayer substrate with the antennas in the upper layer and the rectifiers in the lower layer and GND in the middle layer. The buck-boost converter is also in the lower layer, and is connected to the output of the rectenna array. We connect the feed point of the antenna and the input port of the rectifier by a through hole which is isolated from the middle layer. The GND in rectifiers and buck-boost converter are all connected to the middle layer by other through holes. Figure 5.14 shows a side-view diagram of the multilayer substrate. In the upper layer, antennas are arranged in an equilateral triangle position with the separation between two near antennas of 75 mm, which is almost 0.61 λ (λ is a free-space wavelength at 2.45 GHz). The rectenna array has two types such as A and B, and the buck-boost converter also has two types such as O-type and I-type. With those different types, there are four different combinations of the rectenna array and the buck-boost converter such as A rectenna array + O-type converter (9AO), A rectenna array + I-type converter (9AI), B rectenna array + O-type converter (4BO) and B rectenna array + I-type converter (4AI), respectively. The rectenna and converter combination is referred to as compact power-receiving device here. The photograph of the 9AO and 4BI power-receiving devices are shown in Fig. 5.15.

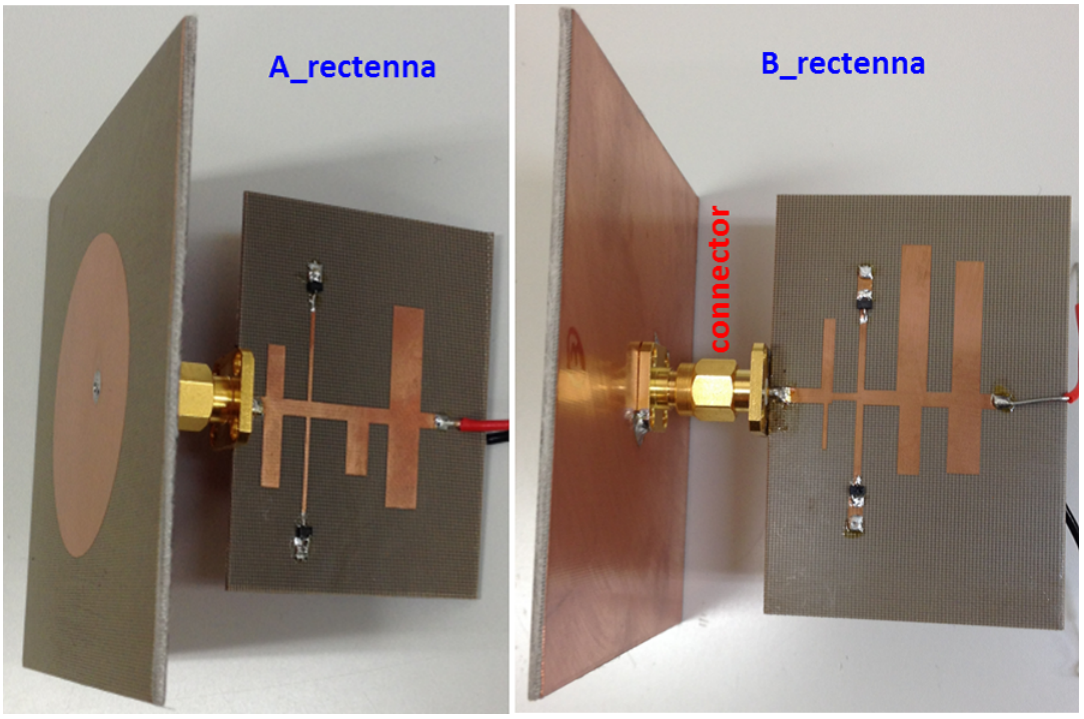


Figure 5.13: Photograph of a single A rectenna and B rectenna.

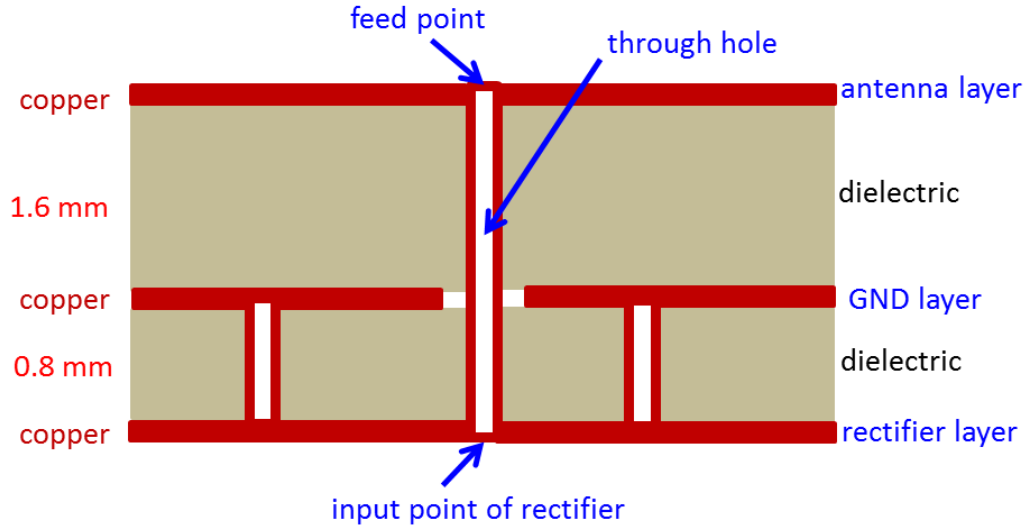


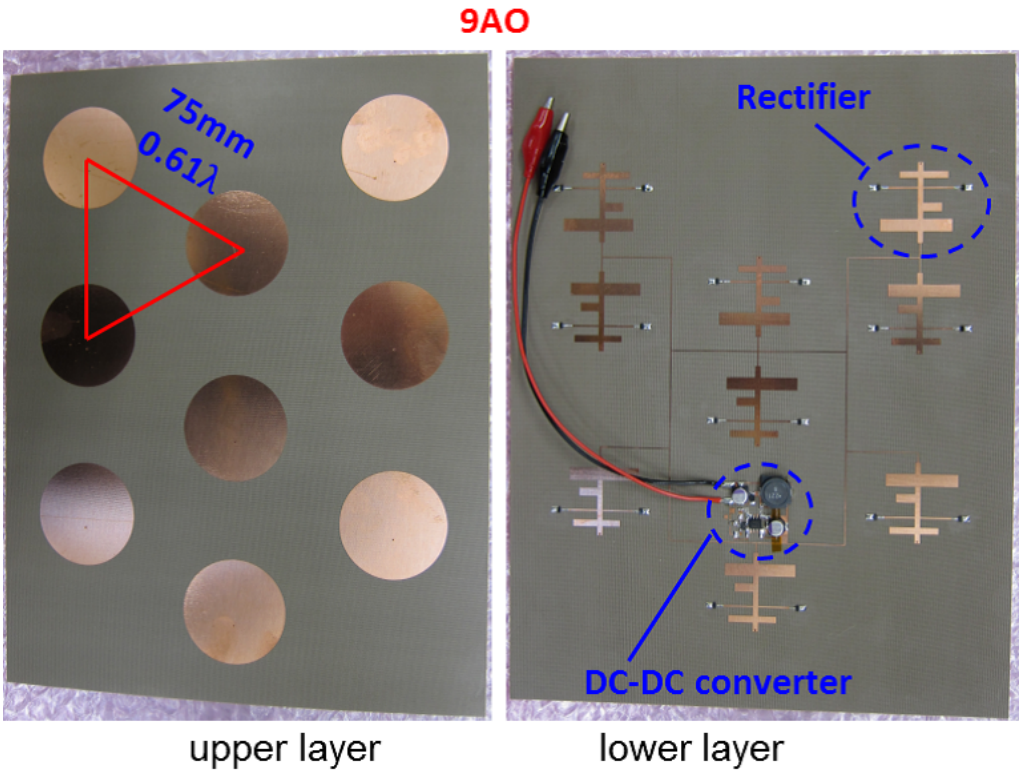
Figure 5.14: Side-view diagram of the multilayer substrate.

5.6 MPT Experiment Using Continuous-wave

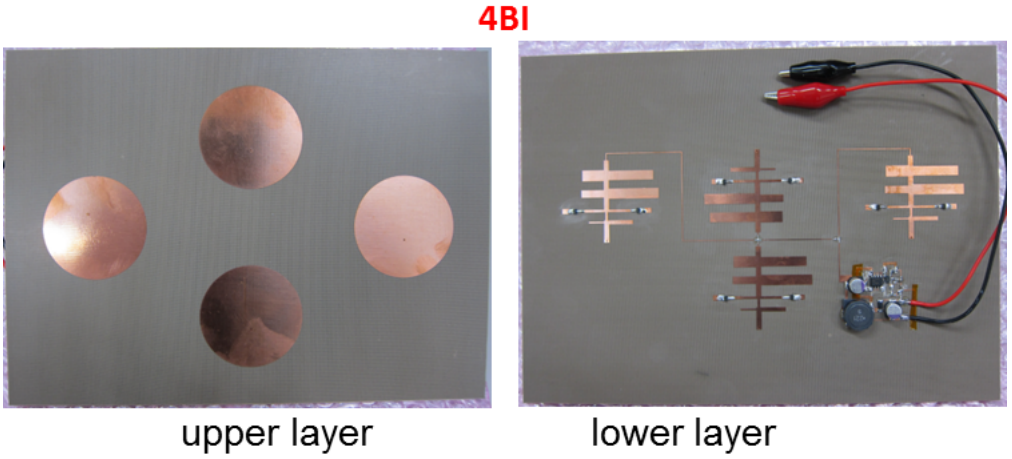
5.6.1 Power Received by Antenna Array

In order to evaluate the overall efficiency of the compact power-receiving device, it is first necessary to measure the power received by the antenna array P_{rall}. Figure 5.16(a) shows a block diagram of the experimental setup used to measure P_{rall} of 9 antennas. It consists of a signal generator (SG), an amplifier, a directional coupler, a horn antenna and the receiving antenna array (without the rectifying circuits or buck-boost converter). The SWH-22 horn antenna of ARA Technologies Inc. we used has an aperture size of 36 mm×25 mm and a gain of 16.3 dBi. According to the antenna far field distance d_{f} calculation, $d_{\text{f}} = 2D^2/\lambda$, where D is the largest dimension of the aperture, the far field distance d_{f} is calculated to be 21 cm. Hence, we set the microwave power transmission distance at 3 m which is much longer than d_{f}.

The power meter measures the input power of the transmission antenna through power sensor A, and measures the received power of each individual antenna one by one through power sensor B. In order to eliminate the influence from the reflection of near antennas, the other eight antennas are all connected to a 50 Ω dummy load as one antenna is being measured. The example of 9 antennas measurement is shown in Fig. 5.16(b). For comparison, the power received by just



(a)



(b)

Figure 5.15: Photograph of the compact rectenna array and buck-boost converter. (a) 9AO power-receiving device. (b) 4BI power-receiving device.

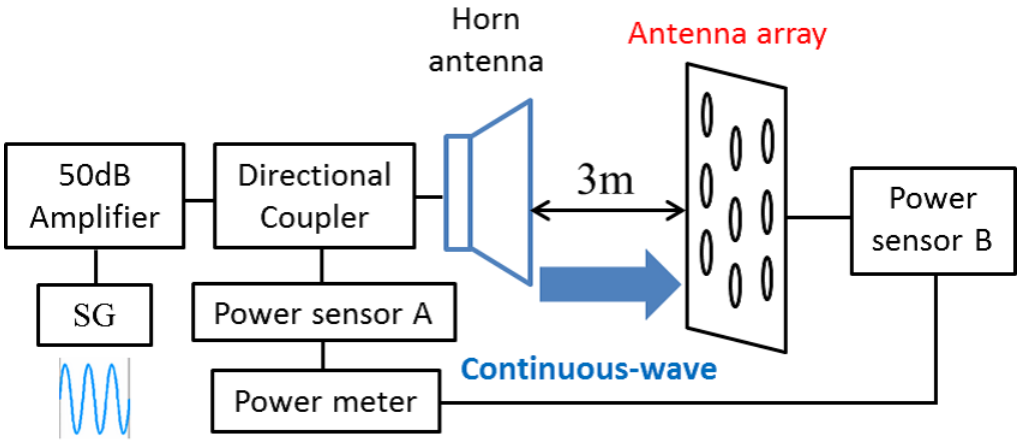
Table 5.5: Relative power received by 9 antennas array and a single antenna vs. power density (P_{den}).

P_{den} (mW/cm^2)	0.25	0.5	0.74	0.98	1.22	1.46	1.67	1.89	2.09
a1	0.879	0.882	0.882	0.881	0.877	0.883	0.881	0.877	0.89
a2	0.828	0.831	0.83	0.833	0.829	0.829	0.828	0.828	0.839
a3	0.738	0.737	0.736	0.736	0.735	0.735	0.735	0.735	0.745
b1	0.691	0.692	0.692	0.691	0.689	0.69	0.691	0.689	0.698
b2	0.774	0.77	0.772	0.771	0.768	0.768	0.772	0.772	0.781
b3	0.955	0.956	0.956	0.957	0.955	0.954	0.956	0.956	0.965
c1	0.856	0.86	0.86	0.859	0.857	0.854	0.858	0.857	0.868
c2	0.843	0.844	0.843	0.846	0.846	0.847	0.844	0.842	0.854
c3	0.731	0.735	0.734	0.736	0.732	0.732	0.73	0.731	0.744

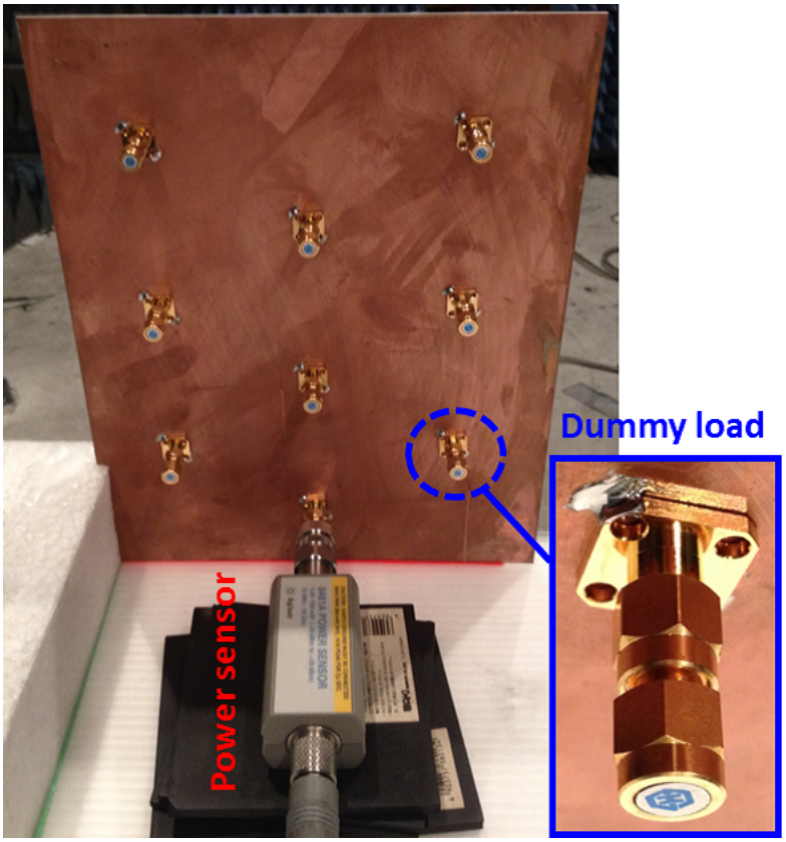
a single antenna, placed at the center position, is also measured, and power received by each antenna of nine antennas are expressed relative to this as shown in Fig. 5.17. These power proportion relation should be constant at different power densities because the gain of antennas are changeless. As shown in Table 5.5, it is obvious that the relative power proportion between a single antenna and antenna array is independent of the power density. The power received by the individual antennas in the array is lower than that received by the single antenna because of overlapping of the effective apertures in the antenna array. Similarly, using the same measurement method, the receiving power of 4 antennas array is also measured. Figure 5.18 shows that the overall power received by the 9 antennas array P_{Aall} and 4 antennas array P_{Ball} in different power densities P_{den}. P_{Aall} increases linearly from 95 to 800 mW and P_{Ball} increases linearly from 45 to 376 mW as the power density is increased from 0.25 to 2.08 mW/cm^2.

5.6.2 Experiment on Driving DC Motor Using Continuous-wave

Figure 5.19 shows a block diagram of the experimental setup used to drive the DC motor by MPT with Continuous-wave power. It consists of the same transmitting



(a)



(b)

Figure 5.16: Experiment setup for measuring the receiving power of the antenna array. (a) Experiment setup. (b) Photograph of the antenna array setup. Power sensor measures the receiving power of each antenna with all other antennas connected to a 50 Ω dummy load.

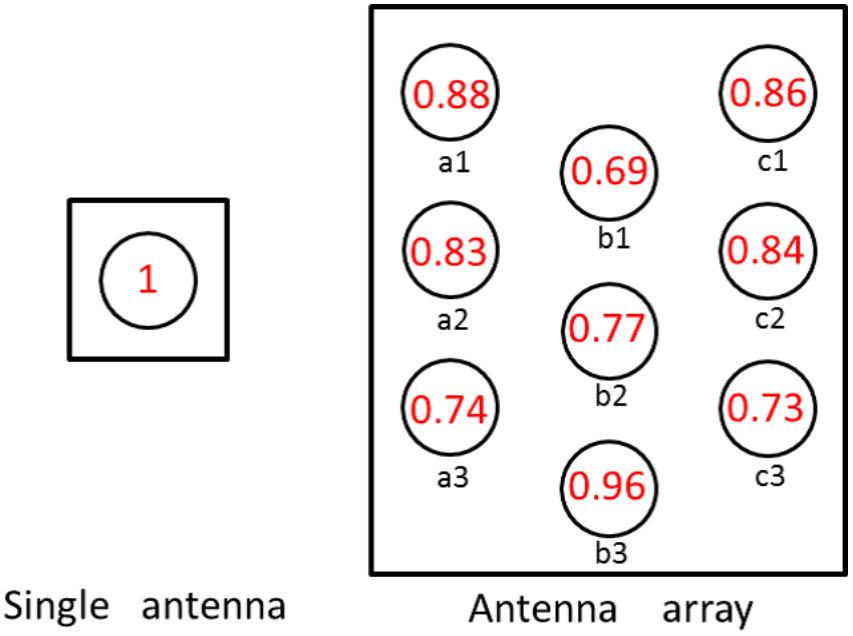


Figure 5.17: Relative power received by 9 antennas array and a single antenna.

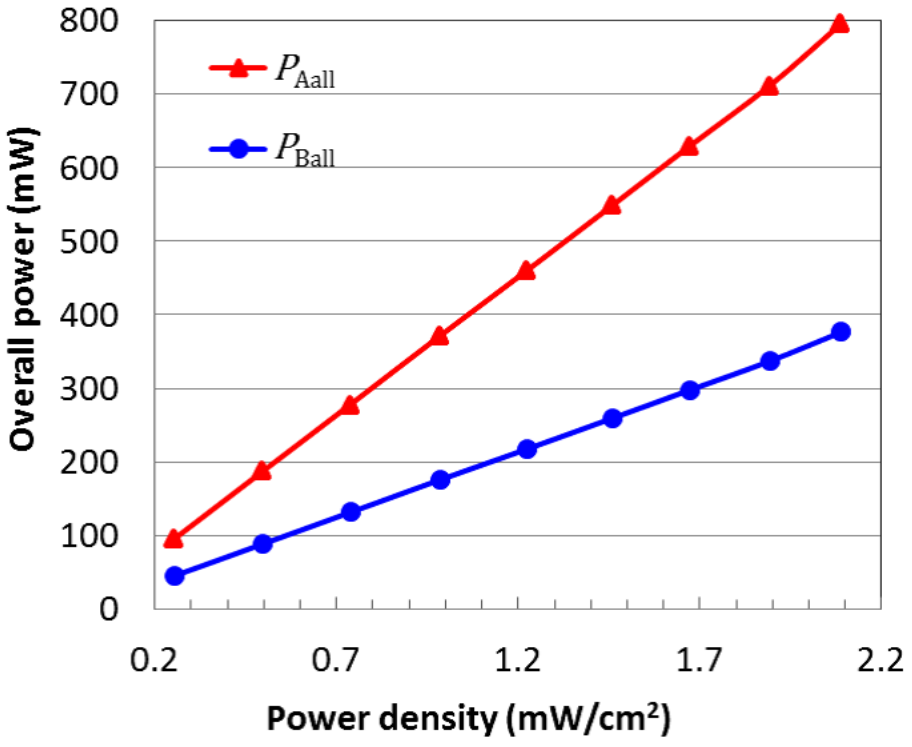


Figure 5.18: Overall powers received by 9 and 4 antennas array such as P_{Aall} and P_{Ball}.

setup as Fig. 5.16, the compact power-receiving device and the DC motor. Of cause, the microwave power transmission distance is also 3 m. The power meter measures the input power of the transmission antenna through a power sensor. An ammeter and a voltmeter measure the input current I_{m} and the input voltage V_{m} of the DC motor, respectively. Firstly, we replace the DC motor with load resistances in Fig. 5.19 to test the compact power-receiving device using CW. Here, the overall efficiency η_{prd} is defined as the ratio of the power output from the buck-boost converter to the power received by the antenna array:

$$\eta_{\mathrm{prd}} = \frac{V_{\mathrm{m}} I_{\mathrm{m}}}{P_{\mathrm{rall}}} \times 100\,\%. \tag{5.2}$$

Figure 5.20(a) shows the overall efficiency η_{prd} of four types power-receiving devices such as 9AO, 9AI, 4BO, 4BI, for load resistances of 10 to 100 Ω with power density P_{den} of 0.98 mW/cm^2. As the load is changing from 30 to 100 Ω, the η_{prd} of the 9AO and 9AI are constant above 70 %, which is slightly higher than that of the 4BO and 4BI of 61 %. This is because that the P_{den} of 0.98 mW/cm^2 reaches the peak efficiency power level for A_rectifier while it is a lower level for B_rectifier according to the results in Chapter 5.4.2. As the load is below 30 Ω, the η_{prd} of four types are low due to the efficiency decrease of the buck-boost converter as shown in Fig. 5.12(a). Figure 5.20(b) shows the output voltage of the power-receiving device, which is also the input voltage of the DC motor V_{m}. Observe that the V_{m} of the 9AO and 9AI are higher than that of the 4BO and 4BI because the former exhibit a higher efficiency than the latter at the same input power and the same load resistance.

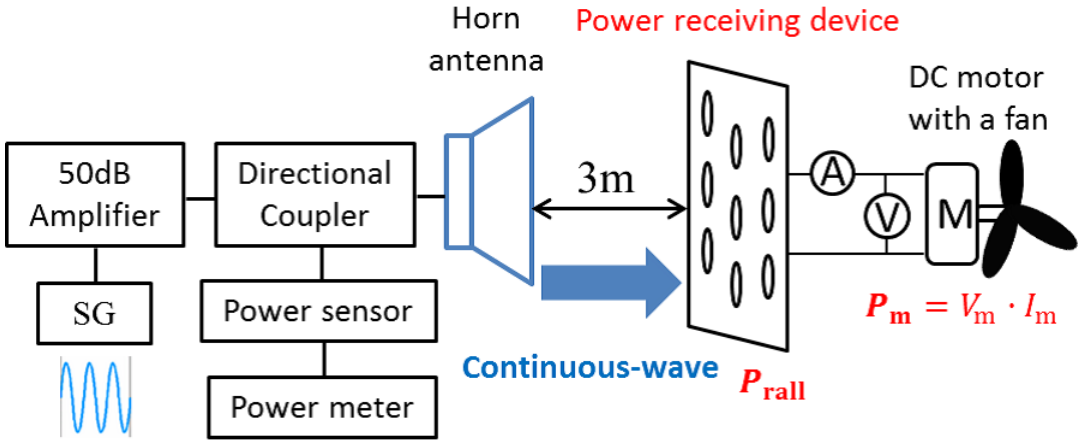
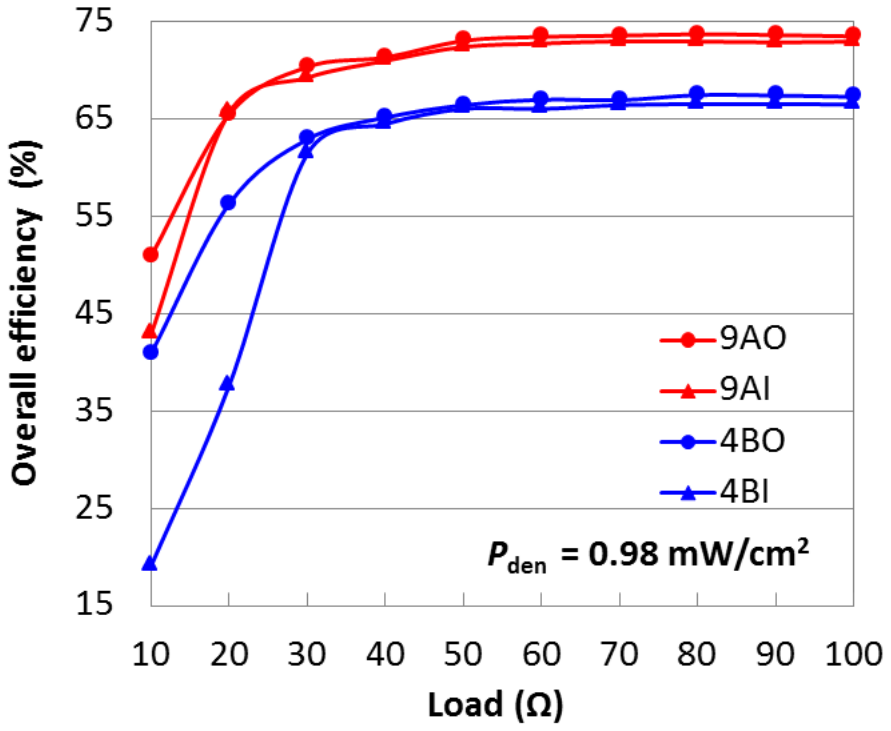


Figure 5.19: Experiment setup for driving DC motor by MPT using CW.



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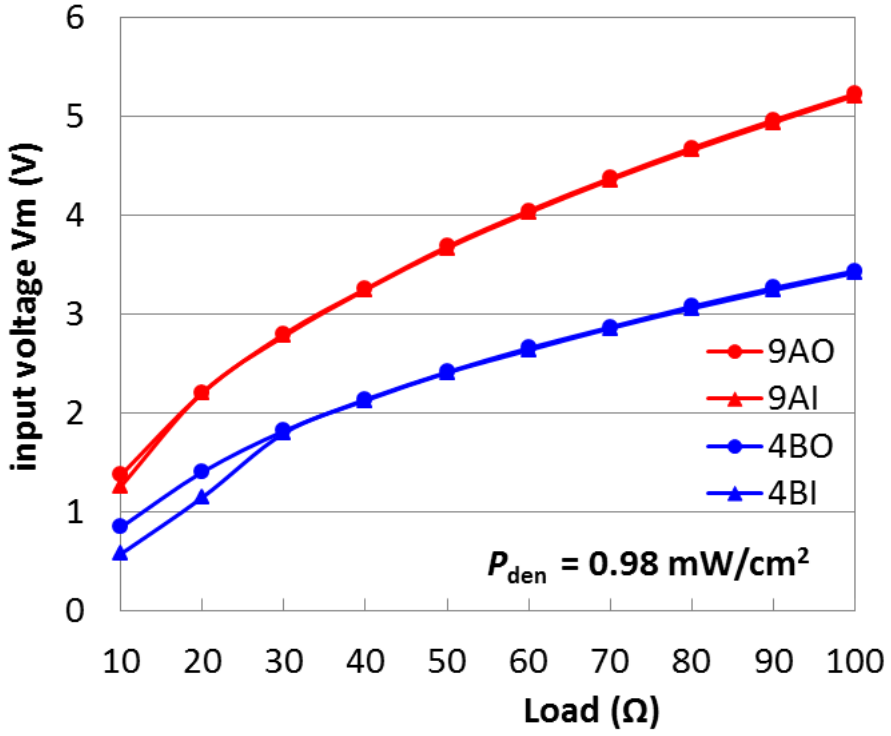
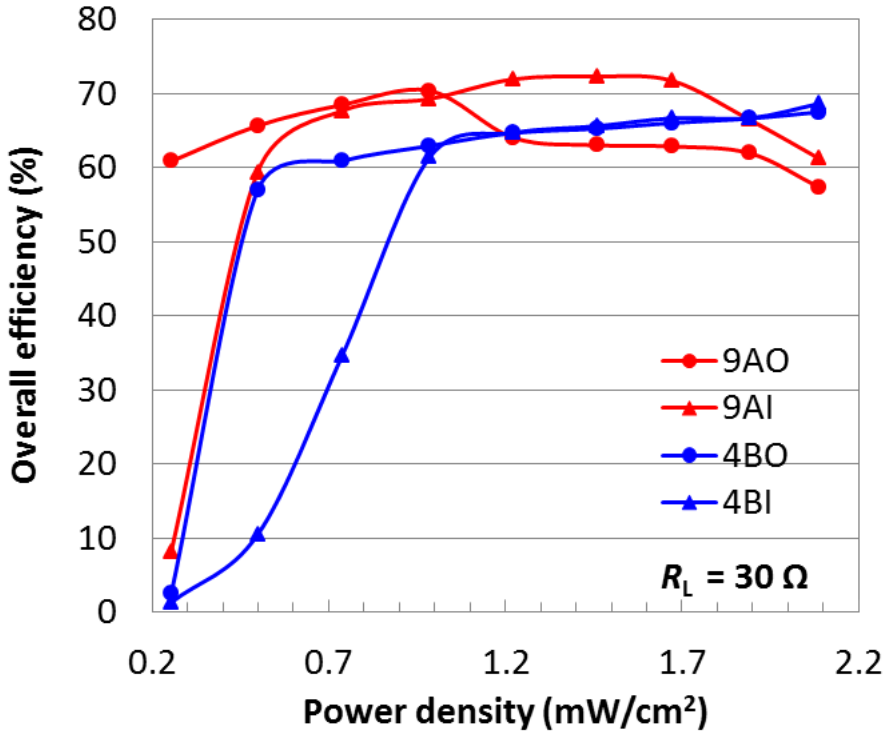


(b)

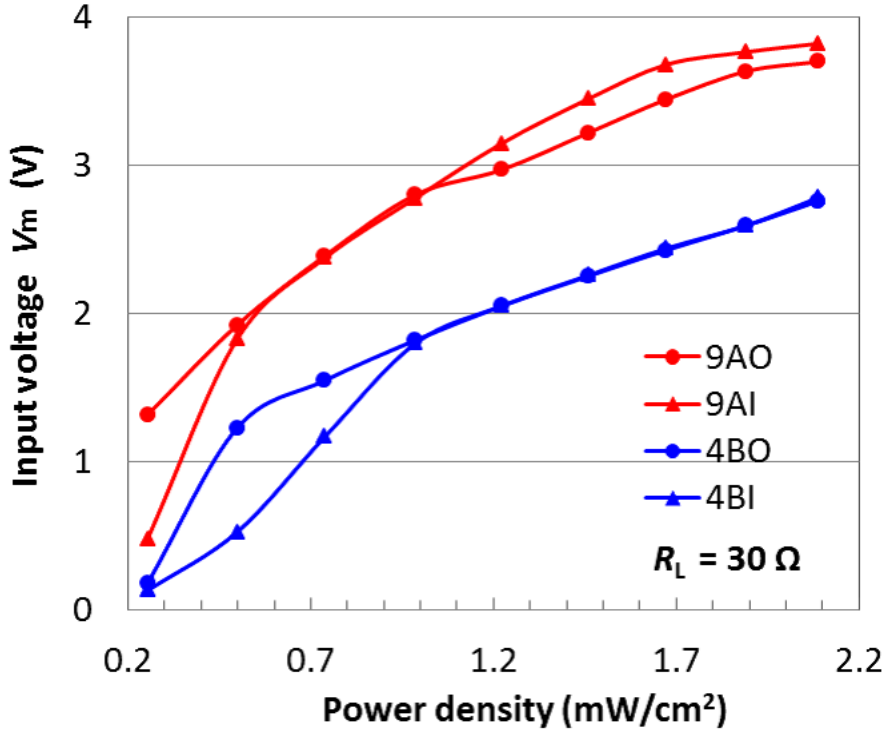
Figure 5.20: Comparison of measured overall efficiency and V_{m} for a fixed power density of 0.98 mW/cm^2. (a) Overall efficiency vs. load resistance. (b) V_{m} vs. load resistance.

Figure 5.21(a) shows the overall efficiency η_{prd} at load resistance of 30 Ω for P_{den} changing from 0.25 to 2.08 mW/cm^2. It is obvious that the η_{prd} of the O-type is higher than that of the I-type both in A_rectifier and B_rectifier for a lower power density level. Compared with 9AI, 9AO performs a higher efficiency when P_{den} is below 0.98 mW/cm^2 while it performs a lower efficiency as P_{den} over 0.98 mW/cm^2. However, compared with 4BI, 4BO performs a higher efficiency when P_{den} is below 0.98 mW/cm^2 while it performs approximately the same efficiency level with the 4BI when P_{den} is over 0.98 mW/cm^2. This is because that the A rectifier obtains a higher output voltage than B rectifier even at the same power density level resulting in efficiency decrease of the O-type converter when the power density is up to a higher level. As for 4BO, the output voltage of the rectifier B is not too large to lead to efficiency decrease of the converter even P_{den} is up to 2.08 mW/cm^2. As shown in Fig. 5.21(b), the V_{m} of the 4BI and 4BO are both below 3 V.

Next, we test the DC motor using MPT in CW power and the measurement overall efficiency results are shown in Fig. 5.22. Similarly, as the power density changing from 0.25 to 2.08 mW/cm^2, the four types overall efficiency are shown in Fig. 5.22(a). Compared with the case of 30 Ω load resistance, the η_{prd} of the 9AO and 9AI almost perform the same variation tendency while the η_{prd} of the 4BO and 4BI perform a little difference in the case of the DC motor. The reason is as following: The input load resistance of the motor R_{m} is increasing as the power density increases which leads to a higher V_{m} and the R_{m} is much larger than 30 Ω in the case of the DC motor. The larger R_{m} leads to V_{m} increase again which results in a higher efficiency of the 4BO because the efficiency of the O-type converter is increasing as the gate voltage increases. The overall efficiency of the 9AO is over 50 % as P_{den} is changing from 0.25 to 2.08 mW/cm^2 while the overall efficiency of the 9AI is over 61 % as P_{den} is changing from 0.5 to 2.08 mW/cm^2. As seen from Fig. 5.22(b), the V_{m} of the 4BO is slight higher than that of the 4BI. Compared with Fig. 5.21(b), V_{m} in Fig. 5.22(b) is higher because the R_{m} is much larger than 30 Ω. In conclusion, 9AO is better for a lower power density level as below 0.98 mW/cm^2 and 9AI is better for a medium power density level as below 1.9 mW/cm^2 and 4BO is better for a higher power density level as over 1.9 mW/cm^2.

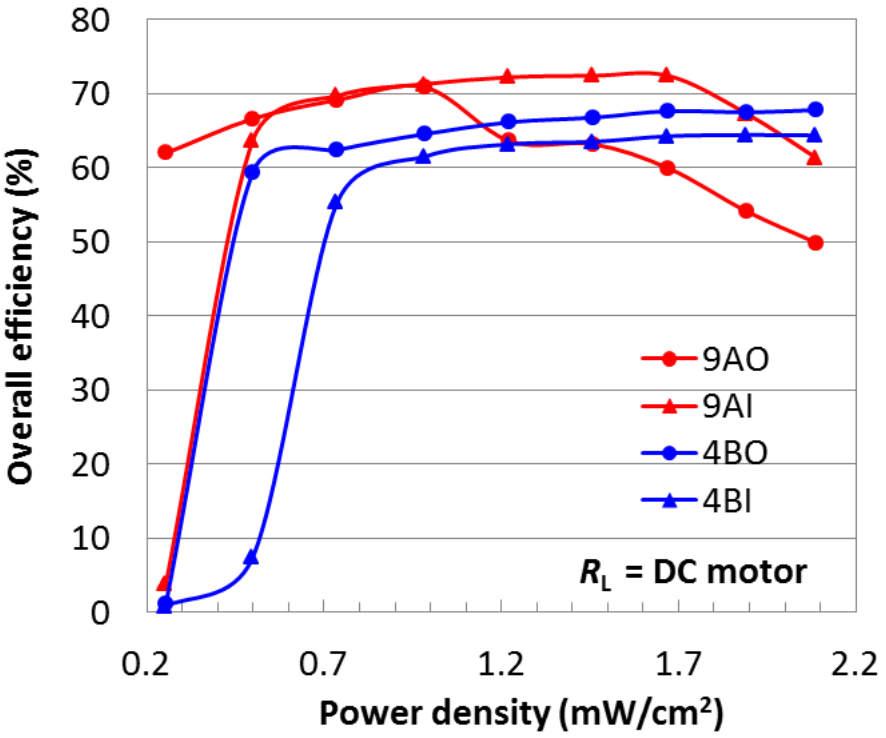


(a)



(b)

Figure 5.21: Comparison of measured overall efficiency and V_m for a fixed load resistance of 30 Ω. (a) Overall efficiency vs. power density. (b) V_m vs. power density.



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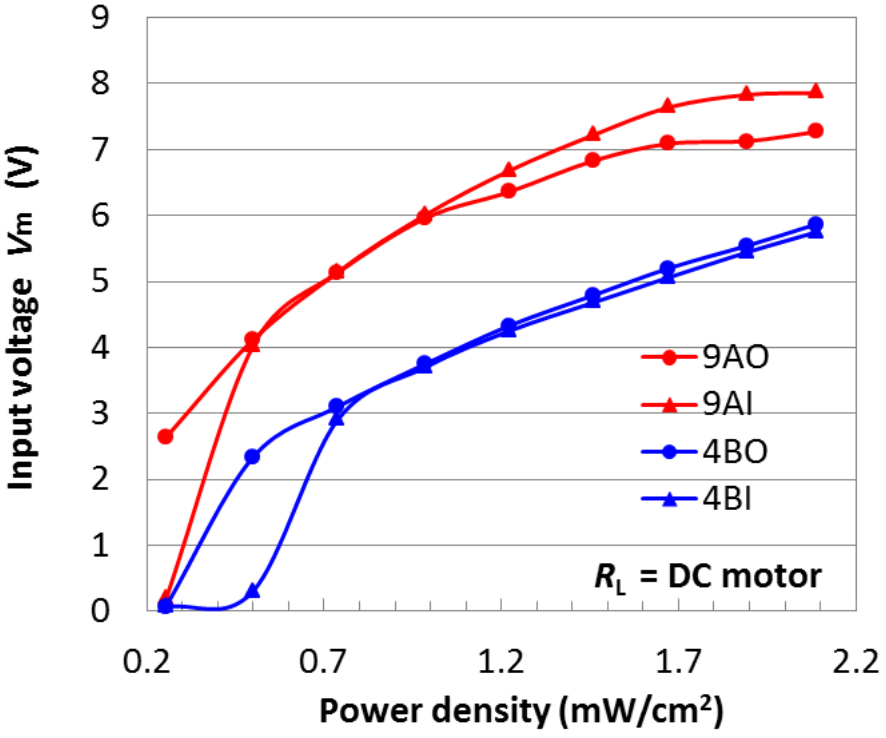


(b)

Figure 5.22: Comparison of measured overall efficiency with DC motor in CW power. (a) Overall efficiency vs. power density. (b) Input voltage of the motor vs. power density.

5.7 MPT Experiment Using Pulsed-wave

Recently, the simultaneous use of MPT and wireless communication has been proposed for ZigBee sensor devices and wireless LAN devices [32]-[34]. The authors proposed that scheduled pulsed-wave power transmission would eliminate or reduce interference between microwave power transmission and communication. Accordingly, we also test the compact power-receiving device using a pulsed-wave power transmission as shown in Fig. 5.23. Similarly, we test it using both load resistances and the DC motor. Here, we set the duty ratio of the pulsed-wave as D_{pw}. Then, the overall efficiency $\eta_{\mathrm{prd}}^{'}$ presents as follows:

$$\eta_{\mathrm{prd}}^{'} = \frac{V_{\mathrm{m}} I_{\mathrm{m}}}{P_{\mathrm{rall}} D_{\mathrm{pw}}} \times 100\,\%. \tag{5.3}$$

Figure 5.24(a) shows the overall efficiency of the 9AO versus the duty ratio of the pulsed-wave when it is loaded by a 30 Ω resistance or the DC motor. The measurement condition is that the power density is 0.98 mW/cm^2 and the pulsed-wave frequency is 25 kHz. Except for a duty ratio of 0.2, similar results are seen for both types of load. A bigger duty ratio leads to a higher input power so that the efficiency of the rectenna is increased. Therefore, the overall efficiency generally increases as the duty ratio is increasing from 0.2 to 1. Furthermore, the overall efficiency for driving the DC motor is over 44 % even though the duty ratio is widely changed from 0.2 to 1. Figure 5.24(b) shows the overall efficiency versus the frequency of the pulsed-wave for driving the DC motor at the power density of 0.98 mW/cm^2 and the duty ratio of 0.5. In this case, the overall efficiency performs a small variation, almost constant at 59 % in the frequency range of 0.33 to 41.7 kHz. As seen from the Eq. (5.3), the overall efficiency is independent of the frequency of the pulsed-wave, which is in agreement with the experiment results. Therefore, the proposed power-receiving device is also suitable for MPT in pulsed-wave power.

5.8 Summary

In this chapter, we conduct several experiments on driving a DC motor using MPT system both in CW and pulsed-wave. Firstly, the background of the study

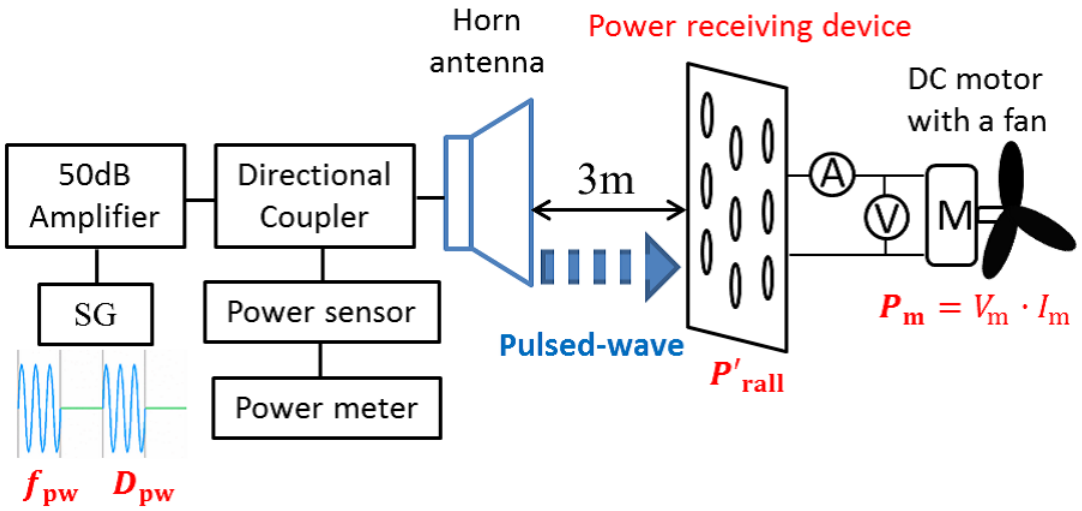
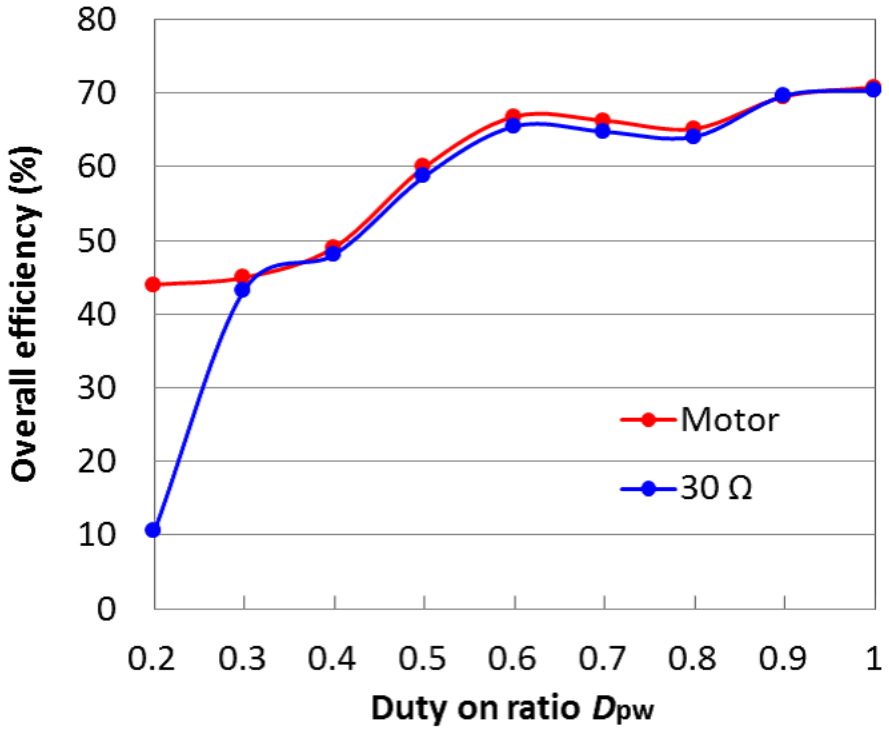


Figure 5.23: Experiment setup for driving DC motor by MPT using pulsed-wave.

that a DC motor is driven by MPT system is introduced. Then, as a test model, we choose a low-power DC motor and the input resistance of the motor is measured. Secondly, we design two antennas with different thicknesses of dielectric substrate of 1.6 and 0.8 mm, and the gain of two antennas are evaluated. We choose the thicker one as the receiving antenna because it exhibits a high gain. Thirdly, to make a compact rectenna array, we design two relatively smaller size rectifiers (A and B) with the measured peak conversion efficiencies of 80.8 % and 80 % at the optimal loads of 300 and 160 Ω, respectively. Fourthly, we design two improved buck-boost converters such as I-type and O-type converters according to the control-pulse oscillator supplied by the input voltage or the output voltage. Fifthly, we compact the antenna array and the rectifier array with the improved buck-boost converter in a multilayer substrate as a power-receiving device. Using those two rectifiers and two converters, there are four combinations of power-receiving devices such as 9AO, 9AI, 4BO, 4BI. Sixthly, we conduct several experiments on driving the DC motor with MPT in CW power using those four combination devices. The overall efficiency of the 9AO is above 60 % in the power density range of 0.25 to 0.98 mW/cm^2 and the overall efficiency of the 9AI is above 60 % in the power density range of 0.5 to 2.08 mW/cm^2. Finally, that the DC motor driven by MPT in pulsed-wave power using 9AO combination device is also tested. In the pulsed-wave case, the overall efficiency is above 44 % in the duty ratio range of 0.2 to 1 for a power density of 0.98 mW/cm^2. Additionally, at a fixed duty ratio of 0.5, the overall efficiency is almost constant at 59 % for the pulsed-wave frequency changing from 0.33 to 41.7 kHz. In conclusion, by



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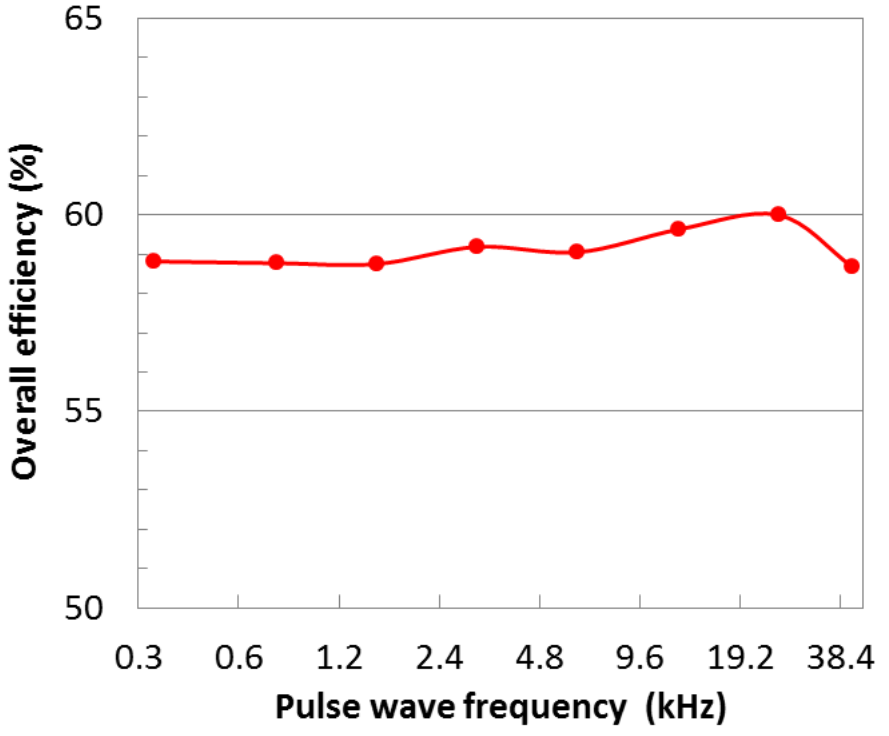

(b)

Figure 5.24: Measured overall efficiency in pulsed-wave mode. (a) Overall efficiency vs. duty ratio of the pulsed-wave. (b) Overall efficiency vs. frequency of the pulsed wave.

both CW and pulsed-wave power, the designed compact power-receiving device performs a relatively high efficiency for driving the DC motor. It is also expected to be valuable for other wireless power transmission applications.

Chapter 6

Conclusion

6.1 Summary and Conclusion

Many researchers are always perplexed by the problem of efficiency-load characteristics in rectenna. This thesis develops a rectenna adapted to ultra-wide load range for microwave power transmission system. We propose a novel resistance conversion DC-DC converter which plays an impedance matching role in rectenna. Connected to this special DC-DC converter, the rectenna seems to operate in a steady load resistance condition despite a varying load resistance connected to the converter. Consequently, when we set the input resistance of the DC-DC converter equal to the optimal load of a general rectenna, the rectenna can keep operating at the peak efficiency point independent of load resistances of the converter.

In Chapter 1, the history of wireless power transmission and microwave power transmission are introduced. An overview of rectenna researches in the world is summarized and the objectives of this thesis are expressed.

In Chapter 2, after giving an overview of efficiency improvement methods in rectenna researches and a comparison of impedance matching methods in WPT studies, we test a PWM controlled boost converter for impedance matching in rectifying circuit. We build a model of a RF-DC-DC circuit which consists of a rectifying circuit and a boost converter, and simulate it using harmonic balance analysis of ADS. Based on the simulation results, the RF-DC-DC circuit obtains a steady high overall efficiency (over 70 %) for a load range from 370 to 1300 Ω, which is wider than that of the rectifier at the same efficiency level. The boost converter

connected to the RF-DC rectifier can limit the maximum reverse voltage of the rectifying diode under breakdown voltage at a wide range of load resistances. The simulation results clearly explain the principle of a DC-DC converter exhibiting as an impedance matching application for rectifying circuit. Therefore, an RF-DC-DC circuit is verified to be available for improving efficiency-load relation of rectenna in this chapter. However, the simulated PWM controlled boost converter just compresses a wide output load range into a narrow input resistance range in some degree. It is insufficient for impedance matching in rectenna.

In Chapter 3, from a new viewpoint, the input/output voltages and input/output resistances relationships of three basic topologies DC-DC converters are discussed. Based on the discussion results, a DCM inverting buck-boost converter, whose input resistance is independent of the load resistance or input power, is chosen for impedance matching in rectenna. We design a negative input voltage buck-boost converter to track the maximum efficiency of the rectifying circuit because the input resistance of the converter can match with the optimal load of the rectifying circuit. As expected, the measured input resistance of the buck-boost converter is approximately 133 Ω despite the changes of the load resistance and input power. Moreover, the measured efficiency of the converter exceeds 80 % for a wide range of loads or input powers. Then we design a rectifying circuit with negative output voltage and obtain a maximum RF-DC conversion efficiency of 79.2 % at an optimal load of 160 Ω with the 2.45 GHz microwave power of 82 mW. Next, we verify that the proposed converter successfully tracks the maximum efficiency of the rectifying circuit. The conversion efficiency of the rectifier is approximately steady at 75 %, despite the load resistance varying from 100 to 5000 Ω for an input microwave power of 82 mW. This means that the maximum power tracking efficiency is over 94 %, compared to the maximum efficiency of 79.2 % at the optimal load. Additionally, the overall efficiency of the RF-DC-DC circuit is approximately steady over 60 %, despite the load resistance changing from 100 to 5000 Ω. According to the experimental results, the proposed RF-DC-DC circuit is an effective MPPT method for low power rectenna. However, the proposed buck-boost converter is difficult for practical applications because of the requirement of an extra DC power supply. Additionally, owing to the loss of the buck-boost converter, the overall efficiency of the RF-DC-DC circuit is much lower than the maximum efficiency of the general rectifying circuit. Therefore, it is necessary to improve the efficiencies both of the rectifying circuit and the converter.

In Chapter 4, firstly, a relatively high-efficiency rectifying circuit is designed and measured. The measured peak efficiency is 82.5 % at the optimal load of 190 Ω for an input RF power of 80 mW. Secondly, we design a self-powered non-inverting DCM buck-boost converter. Compared with the buck-boost converter designed in Chapter 3, it is much superior in capability of handling variable load. Based on the simulation and experimental results, the input resistance of the converter is almost constant at 190 Ω despite the changes of load resistances and input powers, which is in accordance with the theoretical analysis. Moreover, the measured efficiency of the converter exceeds 83 % for a wide load range of 200 Ω to 10 kΩ, and the input voltage is boosted by the buck-booster converter. Finally, we conduct an experiment on RF-DC-DC circuit consisting of the self-powered buck-boost converter and the rectifier. The conversion efficiency of the rectifier is approximately steady at 80 % despite the load resistance varying from 200 Ω to 10 kΩ for an input microwave power of 80 mW. Compared with the peak efficiency of 82.5 % at the optimal load, the maximum power tracking efficiency is over 97 %. Furthermore, the overall efficiency of the RF-DC-DC circuit is constant over 66 % in this ultra-wide load range of 200 Ω to 10 kΩ (a ratio of 1:50). On the other hand, for a fixed load of 800 Ω, the overall efficiency of the RF-DC-DC circuit is also constant over 66 % for the input power of 40 to 120 mW. According to the experimental results, the proposed self-powered RF-DC-DC circuit exhibits a high efficiency in an ultra-wide load range.

In Chapter 5, we conduct several experiments on driving a DC motor using MPT system both in CW and pulsed-wave. Firstly, the background of the study that a DC motor is driven by MPT system is introduced. Then, as a test model, we choose a low-power DC motor and measure the input resistance of it. Secondly, we design two types antenna with dielectric substrate's thickness of 1.6 and 0.8 mm and evaluate them. We choose the thicker one as the receiving antenna because it exhibits a high gain. Thirdly, to make a compact rectenna array, we design two relatively smaller size rectifiers (A and B) with the measured peak conversion efficiencies of 80.8 % and 80 % at the optimal loads of 300 and 160 Ω, respectively. Fourthly, we design two improved buck-boost converters such as I-type and O-type according to the control-pulse oscillator supplied by the input voltage or the output voltage. Fifthly, we compact antenna array and rectifier array with the improved buck-boost converter in a multilayer substrate as a power-receiving device. Using those two rectifiers and two converters, there are four combinations of power-receiving devices such as 9AO, 9AI, 4BO, 4BI. Sixthly, we conduct several

experiments on driving the DC motor with MPT in CW power using those four combination devices. The overall efficiency of the 9AO is above 60 % in the power density range of 0.25 to 0.98 mW/cm^2 and the overall efficiency of the 9AI is above 60 % in the power density range of 0.5 to 2.08 mW/cm^2. Finally, we also test the DC motor driven by MPT in pulsed-wave power using 9AO combination device. In the pulsed-wave case, the overall efficiency is above 44 % in the duty ratio range of 0.2 to 1 for a power density of 0.98 mW/cm^2. Additionally, at a fixed duty ratio of 0.5, the overall efficiency is almost constant at 59 % for the pulsed-wave frequency changing from 0.33 to 41.7 kHz. In conclusion, by both CW and pulsed-wave power, the designed compact power-receiving device performs a relatively high efficiency for driving a load resistance variable device such as a DC motor.

6.2 Thesis Contributions

The contributions of this thesis can be summarized as follows: (1) A relatively high-efficiency rectifier for low-power (80 mW) is developed in Chapter 4. (2) A relatively small size and high-efficiency rectifier for low-power (50 mW) is developed in Chapter 5. (3) The input/output resistance relationships of three topologies DC-DC converters in both CCM and DCM are discussed in Chapter 3, which is expected to be a valuable reference for impedance matching in WPT system. (4) A self-powered buck-boost converter as a resistance conversion circuit whose input resistance is constant and independent of load resistances or input powers. It has been successfully used in rectenna for maximum power point tracking in Chapter 4. It is also expected to be used in other WPT system and energy harvesting system because its constant input resistance characteristics are independent of the operating frequency. (5) A rectenna adapted to pulsed-wave MPT system is firstly discussed here in Chapter 5 and it is expected to be used in coexistent power transmission and communication system. (6) A power-receiving device is developed to effectively drive a DC motor by MPT in Chapter 5, which will be a good reference for researches on airplane or robot powered by WPT system.

6.3 Suggestion for Future Work

This thesis has solved several issues in MPT system. However, a lot work remains to be done and some of the possible future research directions are suggested as follows: (1) In order to effectively simulate the rectifier, a precise HF diode model in ADS or other simulator such as microwave office is required. (2) Because of the supplied voltage limitation from the control-pulse circuit, the input RF power of the RF-DC-DC circuit is limited in the power range of 30 to 140 mW. It is expected to be improved for a much more wide power range. (3) It is necessary to develop a high efficiency (up to 95 %) of the buck-boost converter which can reduce the loss caused by the converter to make the overall efficiency of the RF-DC-DC circuit close to the peak efficiency of the general rectifying circuit. (4) A rectenna, adapted to much small load resistance such as the range of 100 mΩ to 10 Ω, is necessary for charging a battery or driving a high power motor. (5) The impedance conversion buck-boost converter is also expected to be used for other WPT system such as inductive coupling and resonant coupling system. (6) RF harvesting technology and multi-method of energy harvesting technologies are expected to be developed for the application of powering sensor network nodes.

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Publication List

Related Publications

Major Publications

1. Huang, Y., Mitani, T., Ishikawa, T. and Shinohara, N., "Experiment on Driving a Low-power DC Motor by Microwave Power Transmission in Continuous-wave and Pulsed-wave," *IEICE Transactions on Electronics* (conditional acceptance).

2. Huang, Y., Shinohara, N. and Mitani, T., "Development of a Rectifying Circuit of Rectenna Adapted to Ultra-Wide Load Range with Self-Powered DC-DC Converter," *IEICE Transactions on Electronics C,* Vol.J97-C, No.9, pp.342-351, Sept. 2014 (in Japanese).

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Proceedings of International Conference

1. Sakaguchi, K., Yamashita, S., Huang, Y., Yamamoto, K., Nishio, T., Morikura, M. and Shinohara, N., "Experiment of Power Supply Method for WLAN Sensor Using Both Energy Harvesting and Microwave Power Transmissions," in *Proc. PowerMEMS2014,* Hyogo, Japan, Nov. 2014.

2. Huang, Y., Shinohara, N. and Mitani, T., "Improve the Efficiency-Load Characteristic of Rectifying Circuit Using a Self-Powered DC-DC Converter," in *Proc. WPTC2014,* Jeju, Korea, May 2014, pp.197-200.

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1. Huang, Y., Ishikawa, T., Shinohara, N. and Mitani, T., "A Study of Driving Low Power DC Motor Using Microwave Power Transfer," in *IEICE Technical Report,* WPT2014-61, Kyoto, Nov. 2014 (in Japanese).

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Other Publications

Major Publications

1. Sakui, M., Huang, Y., Amei, K. and Ohji, T., "A Method of 2 Pulses Switching for Single-Phase Voltage-Double PFC Converter with Partial Switching Circuit," *IEEJ Transactions on Industry Applications,* Vol.131, No.6, pp.862-863, June 2011 (in Japanese).

Proceedings of Domestic Conference

1. Sakui, M., Huang, Y., Amei, K. and Ohji, T., "A Method of 2 Pulses Switching for Single-Phase Voltage-Double PFC Converter with Partial Switching Circuit," in *The 28th National Conference of the Institute of Electrical Installation Engineers of Japan,* Shinjuku, Aug. 2010 (in Japanese).

Awards

1. 「2013 KDDI Scholarship for International Students」
2. 「2014 Chinese Government Award for Outstanding Self-financed Students Abroad」

Appendix A

Single Shunt Rectifying Circuit

Figure A.1 shows a block diagram of a general rectenna which usually contains an antenna, an input filter, a rectifying component, an output filter and a load. A rectifying circuit is a rectenna circuit without antennas. Many researchers usually discuss the antenna and the rectifying circuit separately. A single shunt rectifying circuit is often used in the rectenna because it performs a high efficiency with a simple circuit structure. The single shunt rectifying circuit has full-wave rectifying capability with one single diode. As shown in Fig. A.2, a single shunt rectifying circuit usually contains an input filter, a rectifying diode, a $\lambda_g/4$ line, a capacitor and a load. Referring to [58], the theory of the single shunt rectifying circuit can be explained as follows.

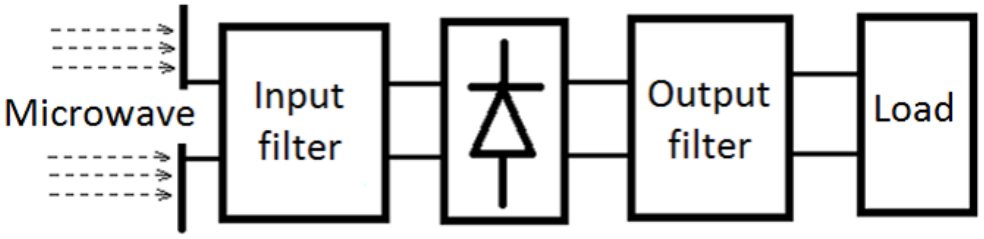

Figure A.1: Block diagram of a general rectenna. It usually contains an antenna, an input filter, a rectifying component, an output filter and a load.

As shown in Fig. A.2, the current $I_1(t)$ is still the high frequency sine wave which can be expressed with Eq. (A.1) and the waveform is shown in Fig. A.3.

$$I_1(t) = I_0 \sin(\omega t). \tag{A.1}$$

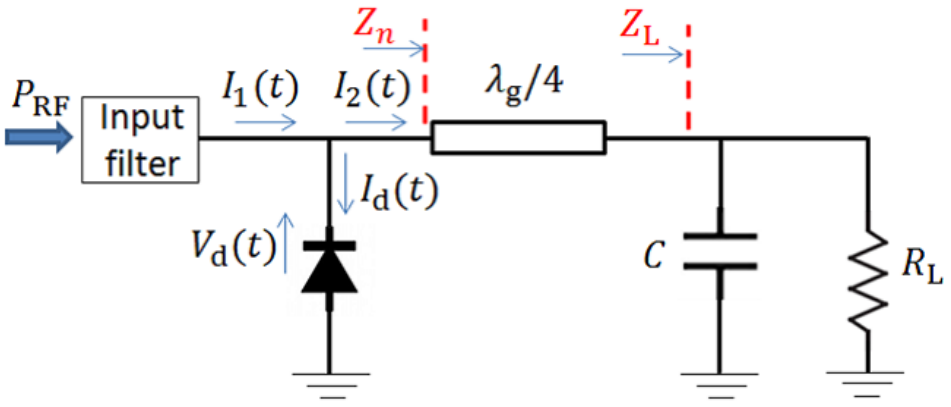

Figure A.2: Block diagram of a single shunt rectifying circuit. It usually contains an input filter, a rectifying diode, a $\lambda_g/4$ line, a capacitor and a load.

Suppose the rectifying diode is ideal, then the diode is open at half period of $0 \leq \omega t < \pi$. It yields the current $I_2(t)$ as following:

$$I_2(t) = I_0 \sin(\omega t), (0 \leq \omega t < \pi). \tag{A.2}$$

Additionally, for rectifying circuit, the impedance of the output filter Z_n seems to be open for odd harmonics and short for even harmonics. It means that only even harmonics current flows through $I_2(t)$. Therefore, $I_2(t)$ satisfies:

$$I_2(\omega t + \pi) = I_2(\omega t). \tag{A.3}$$

Combine Eqs. (A.2) and (A.3), $I_2(t)$ can be rewritten as following:

$$I_2(t) = I_0 |\sin(\omega t)|, (0 \leq \omega t < 2\pi), \tag{A.4}$$

and the waveform of $I_2(t)$ is shown in Fig. A.4 which is a full-wave rectified waveform.

As for the diode current I_d, it satisfies $I_d = I_1 - I_2$. Then:

$$I_d(t) = I_0 \sin(\omega t) - I_0 |\sin(\omega t)|, \tag{A.5}$$

which indicates that the diode current I_d is a half-wave rectified waveform. As for the diode voltage V_d, it is 0 when the diode is on. When diode is off, the voltage applied on the diode includes the DC voltage at the load port (V_L) and the odd harmonics components. Therefore, the diode voltage V_d can be written as:

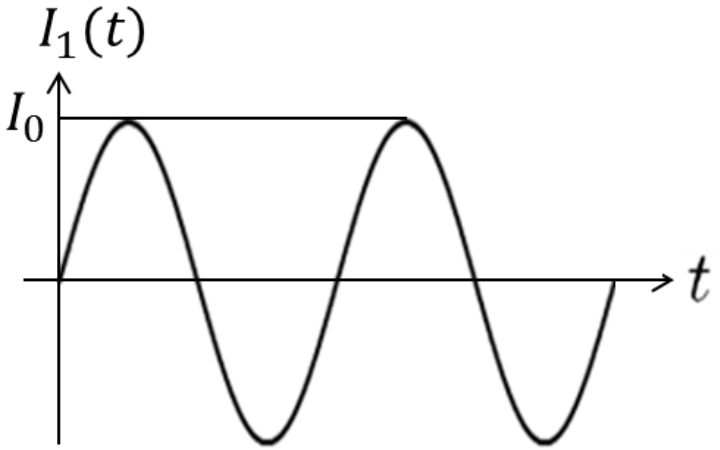

Figure A.3: Current waveform of I_1.

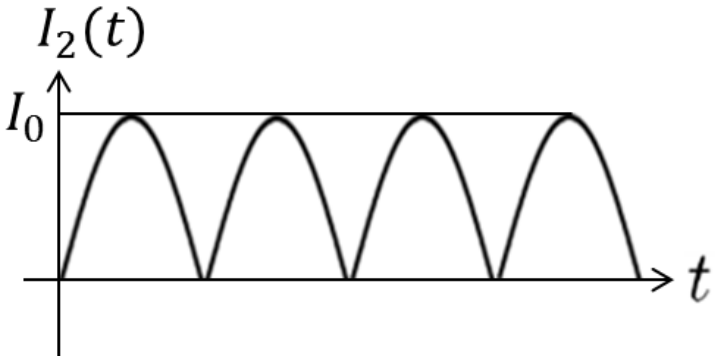

Figure A.4: Current waveform of I_2.

$$V_{\mathrm{d}}(t) = \begin{cases} 2V_{\mathrm{L}}, (0 \leq \omega t < \pi) \\ 0, (\pi \leq \omega t < 2\pi) \end{cases} \tag{A.6}$$

Figure A.5 shows the waveform of V_{d} and I_{d}.

As shown in Fig. A.2, the impedance of Z_{L} and Z_{n} can be written as following:

$$Z_{\mathrm{L}} = \frac{1}{j\omega C + \frac{1}{R_{\mathrm{L}}}}, \tag{A.7}$$

$$Z_{\mathrm{n}} = \frac{Z_{\mathrm{L}} + jZ_0\tan(\beta_{\mathrm{n}}\lambda_{\mathrm{g}}/4)}{Z_0 + jZ_{\mathrm{L}}\tan(\beta_{\mathrm{n}}\lambda_{\mathrm{g}}/4)} Z_0. \tag{A.8}$$

where Z_0 is the characteristic impedance and β_{n} is the phase velocity. ωC is close to infinity as capacitance C is large enough for high frequency. As a result, Z_{L}

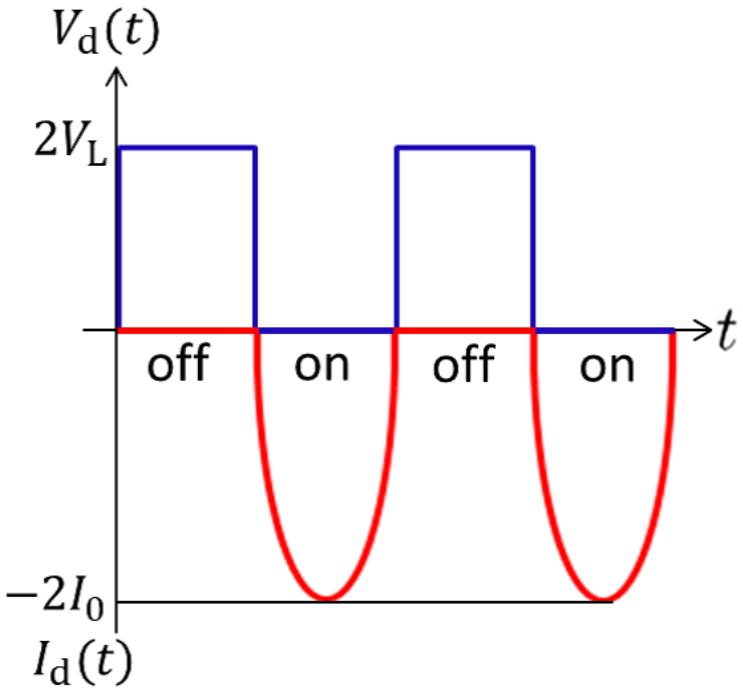

Figure A.5: Voltage and current waveforms of the diode: V_d, I_d.

is close to 0 when ωC is close to infinity. In this case, the impedance Z_n can be rewritten as follows:

$$Z_n = \begin{cases} Z_L, \text{close to 0 (even harmonics)} \\ \frac{Z_0^2}{Z_L}, \text{close to } \infty \text{ (odd harmonics)}, \end{cases} \tag{A.9}$$

which indicates that the impedance of the output filter Z_n is open for odd harmonics and is 0 for even harmonics. Thus, the single shunt rectifying circuit is similar to the class-F amplifier in addressing harmonics problems. Consequently, the single shunt rectifying circuit is also named as class-F rectifier.

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ELECTROMAGNETIC HARVESTING TO POWER ENERGY MANAGEMENT SENSORS IN THE BUILT ENVIRONMENT

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ELECTROMAGNETIC HARVESTING TO POWER ENERGY MANAGEMENT SENSORS IN THE BUILT ENVIRONMENT

by

Evans Sordiashie

A THESIS

Presented to the Faculty of

The Graduate College at the University of Nebraska

In Partial Fulfillment of Requirements

For the Degree of Master of Science

Major: Architectural Engineering

Under the Supervision of Professor Mahmoud Alahmad

Lincoln, Nebraska

May, 2012

ELECTROMAGNETIC HARVESTING TO POWER ENERGY MANAGEMENT SENSORS IN THE BUILT ENVIRONMENT

Evans Sordiashie, M.S.

University of Nebraska, 2012

Adviser: Mahmoud Alahmad

Recently, a growing body of scholarly work in the field of energy conservation is focusing on the implementation of energy management sensors in the power distribution system. Since most of these sensors are either battery operated or hardwired to the existing power distribution system, their use comes with major drawbacks. Battery maintenance and associated labor costs can make implementing sensors very expensive. Although hardwired sensors require very little post-installation maintenance, any energy savings they may procure is offset by the amount of energy expended during the course of the sensors normal operation. In response to these energy challenges, this thesis proposes an electromagnetic harvesting device that powers such sensor technologies by scavenging low electromagnetic field energy. First, the quantities of magnetic flux densities around common current carrying conductors in the built environment are estimated according to the equivalent amount of power that can be generated. Then, a prototype of the harvesting device for scavenging low magnetic flux is designed and developed. Finally, the device is evaluated for real-world implementation using a novel prototype board.

ACKNOWLEDGEMENT

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I would also like to extend my appreciation to my research colleague, Timothy Wisnieski for his continued support throughout this research.

DEDICATION

I would like to dedicate this work to my wonderful family who are my inspiration, my happiness, and my source of energy.

GRANT INFORMATION

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Chapter 1

INTRODUCTION

The efficient management of our energy needs in the built environment requires that we carefully account for each trickle of energy we consume. Recently, engineers have been using efficient energy management sensors as a way to monitor energy and accomplish energy control tasks. Unfortunately, these senor technologies use energy in the performance of such tasks. This thesis focuses on electromagnetic harvesting as a way to power wireless energy monitoring sensors in the built environment. Energy can be saved if sensors are powered by free electric energy scavenged from the built environment. This research looks into creating an effective energy management system that both utilizes an ambient energy source and reduces costs through low maintenance requirements.

1.1 Motivation

This chapter presents the motivation for this research as well as an overview of the various chapters of the research. Since the oil crisis of the 1970s, novel ways to conserve energy have continued to be a top research priority for energy scholars in the United States. Population growth coupled with an insatiable quest for newer electronic technologies has led to a demand for electricity that will surely continue to grow. Unfortunately, it is not yet clear if we are prepared to keep up with this increasing demand in the future. The World Energy Outlook estimates a 76% increase in electricity

use between the years of 2007 to 2030 with an average yearly growth rate of 2.5% (1). However, the demand for electricity has exceeded power transmission growth by 25% since 1982 (2). This is very alarming considering the fact that some 2 billion people currently have no access to electricity (1). Moreover, about 20% of the total CO_2 emissions in the U.S. result from electricity generation alone (3). Figures 1.1 and 1.2 give the U.S. electricity generation by source and equivalent CO_2 emissions from fossil fuels. Since just a little over 13% of our electricity needs are currently generated from renewable energy sources, carbon emissions will continue to endanger the health of the planet for years to come.

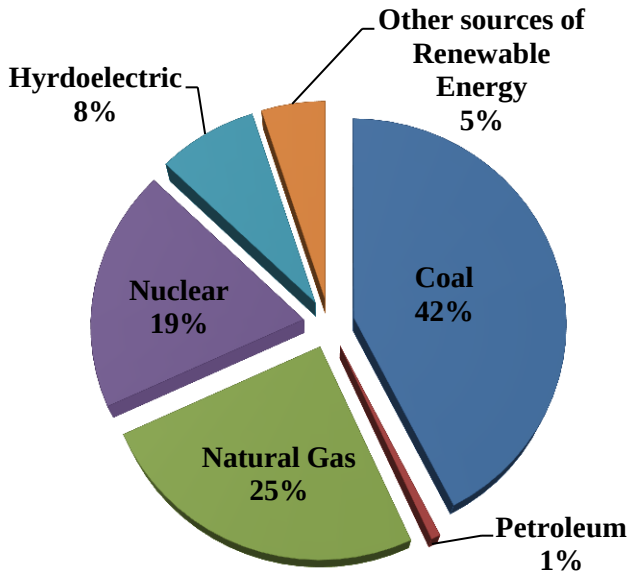


Figure 1.1: US Net Electricity Generation by Energy Source for 2011

Jan, (Source: Energy Information Administration (EIA), Electric Power Monthly, January 30, 2012, Table 1.1), (4).

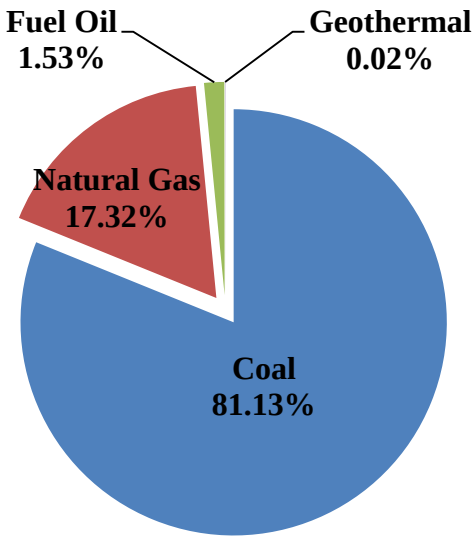


Figure 1.2: CO_2 Emissions from Fossil Fuels Due to Electricity Generation in 2009 (3).

While it should be noted that some scholars have found the relationship between global warming and carbon emissions debatable (5), (6), (7), it is well agreed that our natural resources are being depleted for energy generation.

1.2 Problem Formulation

As part of the effort to curb the rate at which our natural resources are being depleted, it is very important that we not only look for ways to generate energy from renewable sources but also to lower our own energy usage. This means efficiently managing our energy usage both at the residential and commercial level as well as in the built environment as a whole. Recent technologies (8), (9), (10) are focusing on monitoring energy at the node level to conserve energy in the built environment. A node

or outlet is any point in the electrical distribution system at which current is drawn to power electric equipment (11). Real-time energy consumption data can be collected by attaching sensor monitoring devices to these nodes. This has been accomplished in a variety of ways in the prior literature. The investigators in (8), for example, attached wireless sensors to every node in the electrical distribution system for energy monitoring and conservation purposes. However, the sensors themselves consume some energy during normal operation.

Generally, sensors are either battery operated, hardwired from the existing electrical system or are both battery operated and hardwired. Such configurations ensure that the sensor continues to operate during power outages. In a hardwired system, some of the conserved energy will in turn be reused by the sensors during operation. This means that more actual energy will be conserved if the sensors do not draw energy from the existing electrical system. The lifespan of batteries in battery operated sensors last about 9 months in duration. Table 1.1 shows the power consumption data for some common energy monitoring sensors.

Table 1.1: Power Consumption Data for Some Common Energy Monitoring Sensors

Wireless Product	Current Consumption			Battery Voltage	Power Consumption
	Transmission mode	Reception mode	Sleep mode		
RCS-S09U Universal ISM Band FSK Transceiver Module (12)	15mA - 26mA	11mA -15mA	0.3 μA	2.2V-3.8V	33mW - 78mW

G-Link 2.4 GHz Wireless Accelerometer Node (Micro Strain) (13)	25mA	25mA	0.5mA	3.7V	92.5mW
IMOTE2 (Crossbow) (14)	33mA	33mA	390 μA	3.2V-4.5V	127.05mW
XBEE Zigbee/802.15. 4 Modules (DigiInternational) (15)	50mA	50mA	10 μA	2.8V-3.4V	155mW
DataBridge wireless I/O modules (16)	37mA - 120mA	37mA - 120mA	< 100uA	2.7V –3.6V	116.55mW - 378mW
Apex and Apex LT Modules (17)	170mA	37mA	5 μA	2.1V-3.6V	105.45mW - 484.5mW
Lt Series Transceiver Module (18)	12mA - 14mA	12mA - 14mA	11.5 μA-20 μA	2.1V-3.6V	34.2mW - 39.9mW
Si4420 Universal ISM Band FSK Transceiver (19)	13mA - 26mA	11mA -15mA	0.3 μA	2.2V-5.4V	41.8mW - 98.8mW

Batteries for battery operated sensors require periodic replacement after they lose all of their energy. Calculations for the total replacement costs must therefore include not only the cost of the batteries but also any additional labor requirements. An average home containing 30 nodes equipped with such sensors will thus be quite expensive to maintain.

As a way to tackle the power demand of these energy management sensors, this thesis proposes the use of electromagnetic energy harvesting to supply their power needs. Figure 1.3 is a conceptual diagram of the proposed energy harvesting method for supplying power to the sensors.

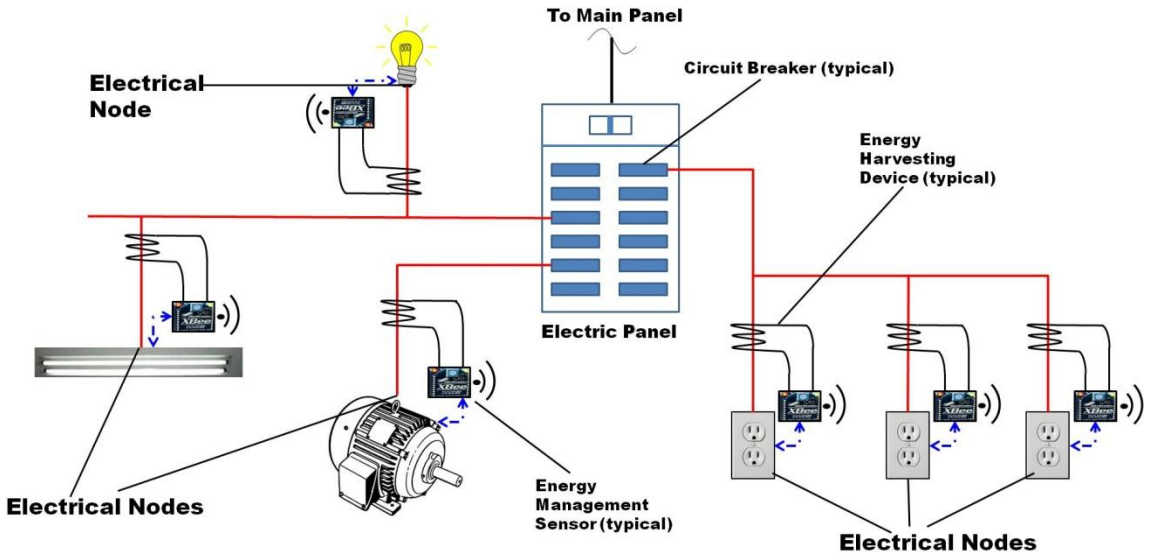


Figure 1.3: Conceptual Diagram of the Energy Harvesting Device in the Electrical System

1.3 Research Goals

One of the goals of this research is to develop a "free" energy source that can power energy management sensors in the built environment. Achievement of this goal will not only increase energy conversation in the built environment but also reduce sensor related maintenance costs. Thus, sensors will eventually be energy neutral and able to operate for an indefinite amount of time.

A second goal of this research is to investigate and determine if electromagnetic energy harvesting is a viable energy source for generating power for other low-powered devices. Electromagnetic harvesting will reduce human dependence on batteries to meet the energy requirements of low-powered devices in the built environment. In fact, it may even make low powered devices autonomous so that they could be employed in hazardous or inaccessible locations in the built environment. It should also be mentioned that one of the related benefits of using the proposed device is that it is an eco-friendly solution, and, with widespread use, may reduce pollution caused by battery disposal

landfills. Last but not least, it is hoped that the proposed work will promote innovations in the built environment that lead to increased energy conservation. This research will open doors to enable the design and manufacturing of future sensor products for electrical energy conservation that incorporate this energy harvesting technology.

1.4 Thesis Overview

The remaining sections of this thesis are broken into 7 major chapters. Chapter 2 gives background information on energy harvesting systems, definitions, and descriptions of common energy harvesting methods, related research and the objectives of this thesis. Chapter 3 gives an overview of electromagnetics and the theory of transformer basics. Chapter 4 describes the research objective, methodology and design. Chapter 5 provides a simulation of and gives experimental data on the proposed energy harvesting device. Chapter 6 summarizes some of the technical challenges involved with use of the device and possible future applications of the device. Finally, Chapter 7 concludes with a review of the information presented in this thesis.

Chapter 2

BACKGROUND

Chapter 2 introduces the definitions, terms, energy harvesting methods and related work that form the foundation for the work presented in this thesis. This chapter will also discuss the proposed work.

2.1 Energy Harvesting System

Energy harvesting systems are made up of three basic components: The energy source, energy scavenger (active component) and external electrical circuit. Figure 2.1 shows a typical energy harvesting system.

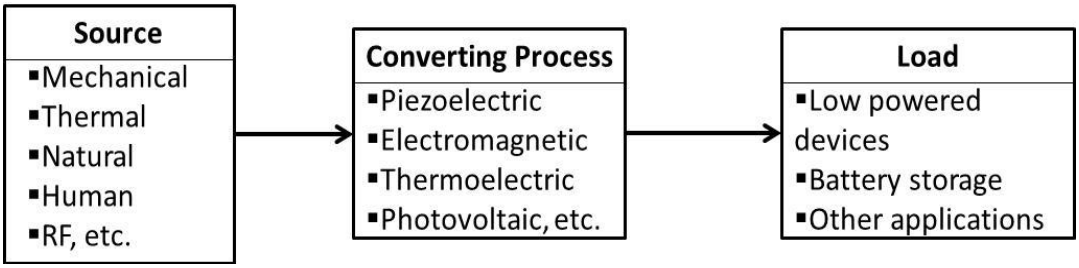


Figure 2.1: Components of Energy Harvesting Systems

In order to harvest useful electrical energy, these three basic components must be in place at all times. The following section gives a summary of common energy sources.

2.1.1 Energy Sources

The most common energy harvesting sources are natural (solar, wind, ocean, etc) energy, mechanical energy, radio frequency (RF), and human (physical) energy (20),

(21). Other types of energy sources include thermoelectricity (temperature variations) and electromagnetic energy.

Sources of natural energy include solar, wind, ocean waves, and water movements. Energy derived from these sources is mostly called renewable energy. Close to 13% (see Chapter 1) of the total electrical energy generated in the United States comes from renewable sources. Of these sources, conventional hydroelectric energy is the leading supplier or provider. The production of hydroelectricity involves using the gravitation pull of water to turn a generator powered turbine. China is currently the world's leading producer of conventional hydroelectricity followed by Canada and the United States, respectively (22).

Recently, solar and wind power are becoming popular sources of energy to harvest. Energy from light (solar power) is directly converted to electrical energy using solar cells (photovoltaic cells). Sunlight is the major source of light for most solar power generation plants. Some common smaller applications of solar power are portable hand calculators, wrist watches and street lights (23). Wind power is generated through the conversion of energy from a fast moving wind source. Wind turbines, which are typically designed with two or three blades mounted on a shaft from a rotor, capture wind energy (through a spinning action) to generate electricity (24). In order to generate large quantities of energy using this source, power companies have to mount hundreds of individual wind turbines over a widely spaced area. Places in which such mass power can be generated are called wind farms. Wind farms may be located in both onshore and offshore areas.

Mechanical energy is created by the motion or vibrations from machines, mechanical stress and strain from high-pressure motors, manufacturing machines, and other related rotational activities (25). In general, harvesting devices for vibrational sources fall in three main categories: electrostatic, piezoelectric and electromagnetic (26). Small, self-powered devices that use mechanical harvesting are known as micro-electromechanical systems (MEMS), (26), (27).

RF energy is another energy harvesting source that can be broadcast radios, public telecommunication services and other dedicated RF devices (28). The energy harvesting device (rectenna) is able to convert the RF energy into direct current (21). RF energy is inversely proportional to distance and therefore drops as the distance from a source is increased.

Humans generate energy that is a combination of mechanical and thermal energy derived from bodily activities. Running, walking, and other movements produce kinetic energy that can be converted into power. In (29) for example, energy generated from walking was used to harvest electrical energy by the piezoelectric method. In the past, hand-cranked radios and other similar devices were powered by human movement. A summary by Starner of the potential energy that can be harvested from humans is shown in Figure 2.2 (30).

Temperature differences between two points can also be used to generate electrical energy. This is commonly known as thermoelectricity. Thermoelectricity is the direct conversion of a temperature difference into electricity. The device used to generate thermoelectricity is called a thermogenerator (21). Thermoelectricity is referred to in some texts as the “Peltier-Seebeck effect” named after Thomas Seebeck who discovered

electric current flow between the junctions of two different metals at different temperatures.

Activity	*Kilocal/hr*	*Watts*
sleeping	70	81
lying quietly	80	93
sitting	100	116
standing at ease	110	128
conversation	110	128
eating meal	110	128
strolling	140	163
driving car	140	163
playing violin or piano	140	163
housekeeping	150	175
carpentry	230	268
hiking, 4 mph	350	407
swimming	500	582
mountain climbing	600	698
long distance run	900	1,048
sprinting	1,400	1,630

Figure 2.2: Summary of Potential Energy That Can Be Harvested from Humans (30)

Electromagnetic fields are becoming a viable source for energy harvesting. While it is not a naturally available source of energy, it does occur around existing power distribution lines. Using Faraday's law of induction, an electromotive force (emf) is induced in an inductor (coil) placed in a magnetic field. Electricity from Faraday's principle is generated by relative movement of the inductor coil in a permanent magnetic field or an inductor placed in a changing magnetic field.

2.1.2 Converting Processes

Ambient energy exists all around us and can be transformed into a useful energy source. This section gives a brief summary of some commonly available energy harvesting methods.

Photovoltaic Harvesting

The photovoltaic (PV) effect is the process through which a photovoltaic cell (solar cell) converts light into electricity. Photons traveling to the surface of photovoltaic cells are absorbed and their energy transferred to make electricity. Solar power is the most common source of light energy harvesting due to its high power density (31). In addition to natural solar light, indoor fluorescent and tungsten lamp illuminations are also potential sources for harvestable light energy. Much less energy can be harvested from indoor environments compared to outdoor environments since less electricity can be generated by PVs with decreasing light irradiance (24).

Thermogenerator

A thermogenerator is a device that generates electricity by converting the temperature difference between two points, a process known as the thermoelectric effect. Thermogenerators are usually attached to bodies or devices which have a comparable temperature difference in relation to their surrounding environment. A good example of the thermoelectric effect is a wrist watch powered by electricity generated from body heat (32).

Piezoelectric Harvester

Quartz crystals have the ability to generate electricity when pressure is applied to them (33). This process is known as the piezoelectric effect or piezoelectricity. In principle, piezoelectricity involves the accumulation of electric charges in solid materials when mechanical stresses are applied to them. Apart from pressure, other kinetic schemes, such as strain, stress, and distortion, can also be applied to piezoelectric materials to generate electricity. Conversely, a mechanical deformation (shrinking or

expansion) is produced when an electric field is applied to piezoelectric materials. Commonly used piezoelectric materials are ceramics. Some useful applications are ceramic resonators, ceramic bandpass filters, ceramic discriminators, ceramic traps, SAW filters, and buzzers (34).

RF Energy Harvester (Rectenna)

The rectenna is the harvesting device used to scavenge power from RF energy (21). It consists of a receiving antenna and a rectifier circuit which converts RF or microwave power into useful DC power (35). The five main components of a rectenna are an antenna, input low-pass filter, rectifying diode, output filter, and resistive load (36). Design and experiments of a high-conversion-efficiency rectenna are found in (36).

Electromagnetic Harvester

From Faraday's law of induction, an induced electromotive force is generated in an inductor (conductor) placed in a magnetic field. This principle is also called electromagnetic induction. The conductor typically takes the form of a coil and electricity is generated by either the relative movement of the coil and the magnet, or via a stationary coil in a changing magnetic field. The converse of this principle is also true. Thus, a magnetic field is generated around a conductor carrying electrical current. The research proposed herein focuses on generating electrical energy by exploiting this combined principle. A magnetic field is generated around current carrying conductors in the built environment. This magnetic field is used in this research to produce electricity for the purposes of supplying energy to energy management sensors. Since the sensors are located close to the electrical nodes, the electromagnetic field generated around the current carrying conductors becomes an ideal source for energy harvesting.

2.1.3 Load

Common use of energy harvesting application is for supplying power to low powered devices. These devices are mostly low powered sensors used in remote or hazardous locations for system monitoring. Table 1.1 in Chapter 1 lists some of these sensors. In harvesting systems where power generated is not constant, battery storage systems are incorporated to store power during peak times to be used later. This is common with photovoltaic systems which typically incorporate batteries for power storage during the day to be used later at night (24). Other common applications of energy harvesting are solar powered calculators and lamps, thermal powered wrist watches, etc. The next section reviews some of the related research in electromagnetic energy harvesting which is the proposed harvesting method of this research.

2.2 Related Work

Several designs for magnetic field harvesting have been proposed in the prior art. Some of the currently available products that harvest energy from electromagnetic fields include the power donut (37) and power line sensor (38), shown in Figure 2.3.

Power donut (24) power line sensor (25)

Figure 2.3: Available Magnetic Field Harvesting Products

The majority of these designs focus on high voltage (HV) transmission lines that have currents in excess of 50A.The authors of (39) propose a microcontroller controlled electromagnetic energy harvesting system for harvesting energy from a single HV transmission line with current between 65A and 130A. The system comprises a toroid magnetic power generator setup, voltage multiplier, DC-DC boost converter and microcontroller. An adaptive algorithm is incorporated into the microcontroller to adjust the duty cycle of the converter so that it achieves maximum output power. The magnetic field energy harvesting setup is able to boost 1V AC and can store power at 58mW for 65A flowing through the HV single conductor.

In (40), D. Pimentel et al. presents power management techniques for energy harvesting systems. They introduce Harvesting Aware Power Management (HAPM) strategies for Energy Harvesting Systems (EHS) that keep devices operating in an energy neutral mode. The authors consider the advantages and disadvantages of three different methods of HAPM: Duty Cycling, Dynamic Voltage, or Frequency Scaling and Maximum Power Point Tracking. The authors conclude that dynamic or adaptive power management is the best choice for energy harvesting systems (2010).

Rohit Moghe et al. (41) study the electric and magnetic field energy harvesting for wireless sensor networks in power system applications. Their work begins with a market survey on the power consumption of a selection of wireless communication devices. Surveys on vibration, solar, and thermal based energy harvesting products are also presented along with information on energy harvesting methods in vibration, thermal, solar EM Wave and Magnetic Field Energy. Theoretical analysis of their electric-field harvesting shows a maximum harvested power of 95.2mW. Experimental testing of a

piezoelectric based energy harvesting device is also conducted revealing that up to 108.24μW can be harvested for a 1000A current carrying conductor. Experimental testing for magnetic field energy harvesting conducted by Rohit Moghe et al. for various core and winding configurations is shown in Table 2.1(41). The maximum power generated is found to be 257mW (2009).

Table 2.1: Magnetic Field Energy Harvesting for Different Core Configurations (41)

Type of Coil	No. of turns	O.C. Voltage at 200A Primary current (V)	O.C. Voltage at 1000A Primary Current (V)	Max. Harvestable Power (mW)
Rogowski Coil	18	0.03	0.16	8
28AWG wire wound on a Wooden Core	200	0.24	1.21	29.8
28AWG wire wound on a hollow semi cylindrical Silicon Steel Core	250	0.37	1.77	210.2
28AWG wire wound on a Flux Concentrator	300	0.50	2.64	257
Flux Concentrator Connected to a Transformer	300	12.5	70.6	225

Rashed H. Bhuiyan et al. propose an energy coupler for scavenging energy from current carrying conductors to power wireless sensors in electrical systems (42). Their device uses a magnetic core with an air gap in between layers. The core material presented in their work is a mu-metal of cylindrical form. Figures 2.4 and 2.5 show the diagrams for the core configuration used in their theoretical analysis (2010).

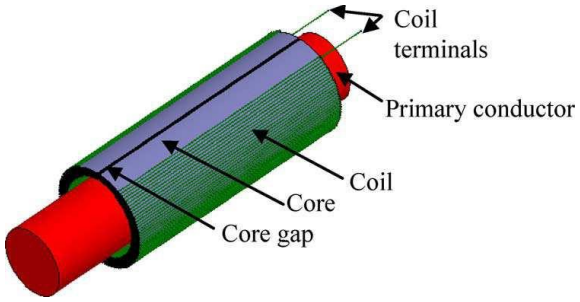


Figure 2.4: Multiturn Coil on a Magnetic Core Around A Current Carrying Conductor (42)

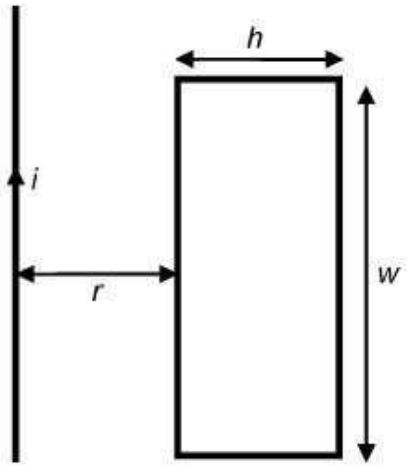


Figure 2.5: Rectangular Coil Near a Current Carrying Conductor (42)

The voltage induction equation is given as (42):

$$V = N\mu_0\mu_r wfI\ln\left(1+\frac{h}{r}\right)$$

Where N – number of turns wound around the magnetic core, μ_0 and μ_r are the permeability of free space and relative permeability of the core, *w* and *h* are the width and height of the core, *f* is the operating frequency, *r* is the distance from the current carrying conductor and *I* is the root mean square (rms) magnitude current in the current carrying conductor. A prototype of the coupler is built and tested in an experiment setup with a board consisting of eight parallel 200 W light bulbs, 120V, 60 Hz supply main (42). Multiple turns of the primary conductor are passed through the opening of the coupler each carrying 13.5A. The energy coupler is shown to be capable of producing 10mW of DC power to a 50 Ω. The authors estimated a higher power to be generated by the coupler with conductors carrying currents in excess of 1000 A (2010).

Except for the work presented in (42), no research to the author's knowledge looks at harvesting electromagnetic energy from low current drawing conductors. Moreover, the research presented in (42) uses multiple primary turns to generate an ample amount of power. The highest voltage induced was approximately 1.5V AC for 5 primary turns and 280 turns of secondary winding for 13.5A current in the primary conductors (42). However, even though the energy coupler in (42) has praiseworthy features such as its ease of use and miniature size, the low current range in the built environment (typical, 1-30A) can reduce it capacity to generate power. This is due to the fact that not enough magnetic flux can be transferred to secondary windings for power generation with the configuration of the proposed energy coupler. Figure 2.6 shows a graph of measured induced voltages from Bhuiyan et al. (42). For practical applications, a single primary turn will typically be used in the energy coupler. From the graph in Figure

2.6 it can be seen that the amount of voltage induced for a single primary turn is a little over 0.2V for 280 turns with a primary current of 13.5A. This makes the coupler presented ideal for high current drawing conductors.

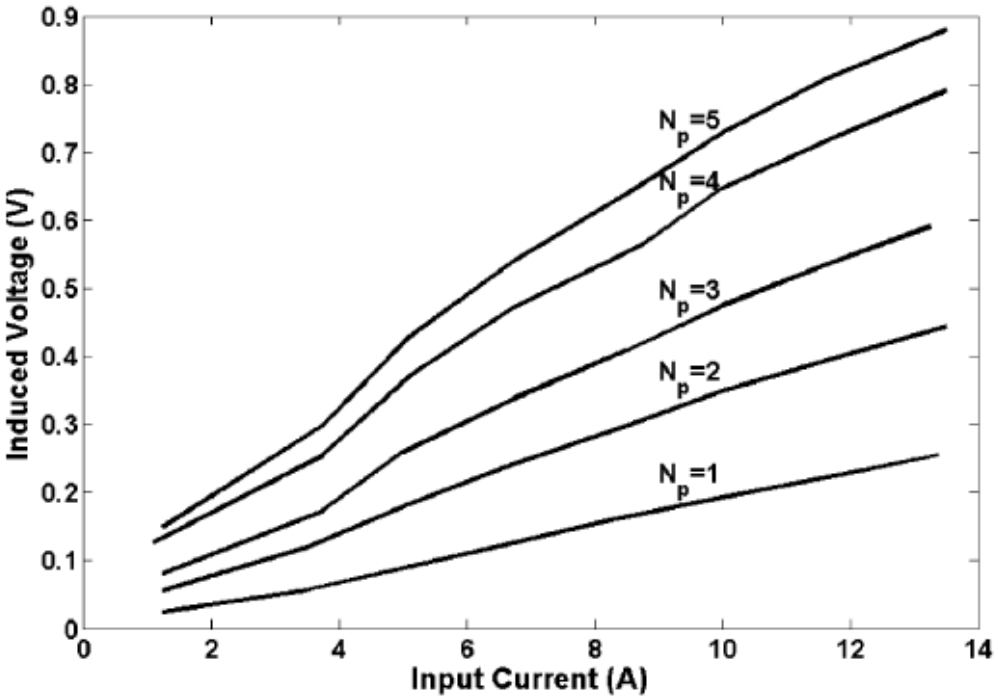


Fig. 11. Measured induced voltages versus currents (I) with number of turns (N_p) of the primary current conductor as a parameter. Parameters: $w = 50\,\mathrm{mm}$, $n = 8, T = 1.25\,\mathrm{mm}, N = 280, l_g = 1\,\mathrm{mm}, d = 0.1016\,\mathrm{mm}, f = 60\,\mathrm{Hz}$, and $\mu_r = 10^5$.

Figure 2.6: Measured Induced Voltages From (42)

The harvesting device proposed in this research uses only a single primary current carrying conductor with multiple secondary winding s from 200 to 500 turns. The core used in the research is an amorphous material designed at a reduced size to better scavenge magnetic energy. The core material has a saturation flux density of 1.56 T (43) and a rectangular cross-sectional area. Theoretical analysis used is based on design of current transformers. In order to effectively harvest ample amounts of energy, the available magnetic flux around the current carrying conductors (1-30A) in the built

environment is theoretically analyzed and simulated. The equivalent voltage that can be induced from this available magnetic flux is also analyzed and simulated for a given magnetic core. Magnetic flux and voltages induced is also simulated for various loads at electrical nodes. Simulation of the cross-sectional area of a magnetic core against the voltage induction is done with variable parameters to include the load consumed at the nodes, the current flowing through the primary conductor, and the proximity of the current carrying conductor to the core. Experimental testing is done with an already available standard amorphous core and the results are then compared with the theoretical simulation.

Since the design of the proposed energy harvesting device is based on the electrical transformer, this next chapter gives an overview of transformer basics and current transformer design.

Chapter 3

AN OVERVIEW OF ELECTROMAGNETICS & TRANSFORMERS

3.1 Introduction

Electromagnetic energy harvesting applies the theory of electromagnetism and principles of electric transformer design basics. Electric current flowing to the nodes where energy management sensors are located generates a magnetic field around the current carrying conductor. This magnetic field serves as the main energy source for the proposed electromagnetic energy harvesting device. To estimate amount of available magnetic field strength around the nodes and the equivalent amount of power that can be generated, this chapter offers a review on electromagnetics and transformer basics.

3.2 Overview of Electromagnetics

Electromagnetics involve the theory and application of electric and magnetic fields. Any electric charge in motion has an associated magnetic field. Maxwell's equations are the set of fundamental equations that govern electromagnetics (44). These equations are combination of Faraday's, Ampere's and Gauss' laws (44). For a complete presentation of electromagnetic theory and derivation of Maxwell equations, the reader is referred to references (44) and (45). To estimate the available magnetic field around current carrying conductors for energy harvesting, this section focuses on Boit-Savart law, Ampere's law and ferromagnetic materials.

3.2.1 Biot-Savart Law

Biot and Savart discovered that a conductor carrying a steady current produces a force on a magnet. They developed an equation that gives the magnetic field at some point in space in terms of the current that produces the field. Figure 3.1 shows an illustration of the Biot-Savart law. Thus for a steady current I flowing through the conductor dl, the magnetic field dH at a point P has the following properties:

- The vector *dH* is perpendicular to both *dl* and to the unit vector $\hat{R}$ directed from the element dl to the point P.
- The magnitude of *dH* is inversely proportional to R^2, where R is the distance from the element dl to the point P.
- The magnitude of *dH* is proportional to the current *I* and to the length *dl* of the element.
- The magnitude of *dH* is proportional to $sin\theta$, where θ is the angle between the vectors *dl* and $\hat{R}$.

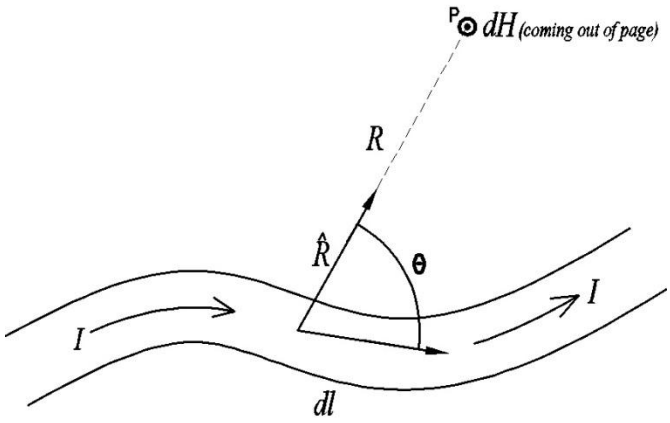


Figure 3.1: Electric Current Relationship with Magnetic Field, Biot-Savart (44)

The differential form of Biot - Savart is given by equations 1, 2:

$$dH = \frac{I}{4\pi} \cdot \frac{dl \times \hat{R}}{R^2} \qquad 1$$

The total magnetic field *H* due to a conductor of finite length *l* carrying a current *I* is,

$$H = \frac{I}{4\pi} \cdot \int_l \frac{dl \times \hat{R}}{R^2} \qquad 2$$

Derivation of magnetic field around a finite length linear conductor

Figure 3.2 shows a conductor of finite length carrying a current in the upward direction.

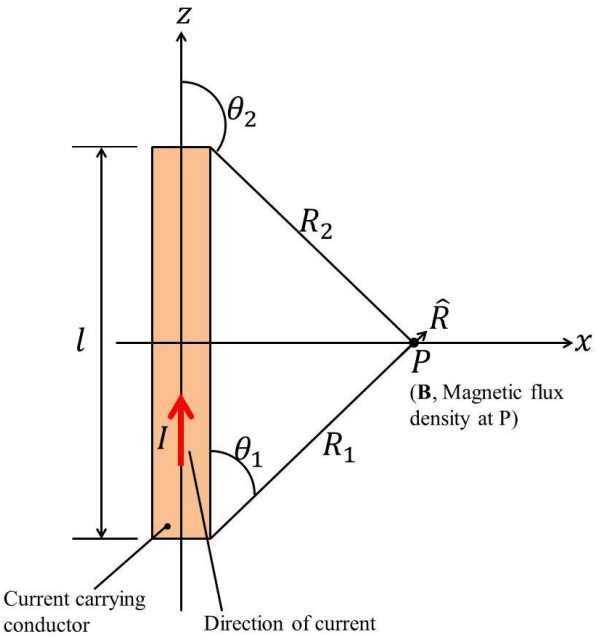


Figure 3.2: Magnetic Field Around Current Carrying Conductor of Finite Length

The integration of equation 2 for a current carrying conductor of length *l* is stated as:

$$H = \frac{I}{4\pi} \cdot \int_{-l/2}^{l/2} \frac{dl \times \hat{R}}{R^2} \qquad 3$$

$$dl = dz\hat{z} \qquad 4$$

$$dl \times \hat{R} = dz(\hat{z} \times \hat{R}) \qquad 5$$

$\hat{z} \times \hat{R} = sin\theta\ \hat{\psi}$, where$\hat{\psi}$ is the unit vector normal to the plane determined by z and R when they are drawn from a common point.

$$H = \hat{\psi}\frac{I}{4\pi}\cdot\int_{-l/2}^{l/2}\frac{sin\theta}{R^2}\,dz \qquad 6$$

Expressing R and z in terms of θ and substituting;

$$R^2 = \frac{r^2}{sin^2\theta},\ z = -r\frac{\cos\theta}{\sin\theta}, dz = \frac{r}{sin^2\theta}d\theta$$

Therefore,

$$H = \hat{\psi}\frac{I}{4\pi}\cdot\int_{\theta_1}^{\theta_2}\frac{\sin^3\theta}{r^2}\cdot\frac{r}{\sin^2\theta}d\theta = \hat{\psi}\frac{I}{4\pi}\cdot\int_{\theta_1}^{\theta_2}\frac{\sin\theta}{r}d\theta \qquad 7$$

$$H = -\frac{I}{4\pi r}(\cos\theta_2 - \cos\theta_1)\,\hat{\psi} \qquad 8$$

Expressing $z = -\frac{l}{2},\ \frac{l}{2}$ in terms of θ_1 and θ_2 ;

$$\cos\theta_1 = \frac{l/2}{\sqrt{r^2+\left(\frac{l}{2}\right)^2}},\ \cos\theta_2 = \frac{-l/2}{\sqrt{r^2+\left(\frac{l}{2}\right)^2}},$$

$$H = \frac{I}{4\pi r}\cdot\frac{2l}{\sqrt{4r^2+l^2}}\hat{\psi} = \frac{Il}{2\pi r\sqrt{4r^2+l^2}}\hat{\psi}(\mathrm{A/m}) \qquad 9$$

Biot – Savart law due to surface and volume distributions are expressed as follows (44):

$$H = \frac{I}{4\pi}\cdot\int_S\frac{J_s\times\hat{R}}{R^2}\,ds \quad \text{(for surface current)} \qquad 10$$

$$H = \frac{I}{4\pi}\cdot\int_V\frac{J\times\hat{R}}{R^2}\,dV \quad \text{(for volume current)} \qquad 11$$

Where, J_s is the surface current density (A/m) and *J* is the volume current density in (A/m^2), *R* is the distance vector between surface or volume and an observation point of interest where magnetic field is being determined (m).

3.2.2 Ampere's Law and Magnetic Flux Density

Ampere's circuital law states that the line integral of a magnetic field H, around a closed path is equal to the current traversing the surface bounded by that path (44), (45). Ampere's law in equation form is given by:

$$\oint_C H \cdot dl = I \qquad 12$$

Where C denotes the contour of close path.

The sign convention for the direction of the contour C path follows the right – hand rule. Thus if the direction of *I* is aligned with the direction of the thumb of the right hand, then the direction of the contour C should be chosen along that of the other four fingers. Therefore, given the direction of current flow in a current carrying conductor, the magnetic field close contour around the conductor will follow the right – hand rule.

The magnetic flux density *B* and field intensity *H* are related by (44), (45);

$$B = \mu H \text{ (T)} \qquad 13$$

Where, μ is the magnetic permeability property of the medium through which the field penetrates. This is usually denoted in text as $\mu_0$for most dielectric materials and metals except for ferromagnetic materials. $\mu_0 = 4\pi \times 10^{-7}$(N/A^2), (44).

For a current carrying wire conductor of finite length *l* shown in Figure 3.2, the magnetic flux density is given by:

$$B = \mu H = \mu \cdot \frac{Il}{2\pi r\sqrt{4r^2+l^2}}\hat{\psi} \text{ (T)} \qquad 14$$

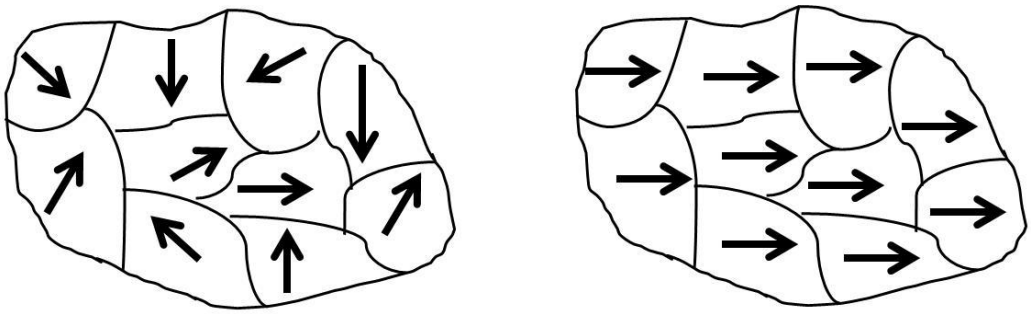
For an infinitely long wire, $l \gg r$

$$B = \frac{\mu I}{2\pi r}\hat{\psi} \text{ (T)} \qquad 15$$

3.2.3 Ferromagnetic Materials

The medium or material for transfer of electromagnetic energy for electricity should exhibit magnetic properties. These materials are called ferromagnetic materials. Ferromagnetic materials are basically materials that have the ability to attract a magnet.

Some common ferromagnetic materials are iron, nickel, cobalt, alnico and an alloy of aluminum-nickel-cobalt. From earlier discussions, the flow of electric charge has an associated magnetic field. The magnetic field comes from the motion of electrons in the atoms of these materials. In addition, each electron spins when in motion. Ferromagnetic materials exhibit a long-range ordering property at the atomic level which causes the unpaired electron spins to line up parallel with each other in a region called a domain (45). Figure 3a shows this alignment phenomenon. Magnetic field is intense in each



a. Domains with magnetic field in material but misaligned with each due to no external magnetic influence.

b. Domains aligned with each other due to external magnetic influence.

Figure 3.3: Domain Alignment in Ferromagnetic Materials

domain but the material as whole is not magnetized due to random orientations of domains (45), (50). However, an external influence of a magnetic field on a ferromagnetic material can cause the domains to line up with each other to cause magnetization. This is illustrated in Figure 3b. The magnetic field is then increased across the material by a factor called its relative permeability. Relative permeability is the ratio of permeability of a material to that of a vacuum at the same magnetic field strength (45). Table 3.1 gives the relative permeabilities of some common ferromagnetic materials.

Table 3.1: Relative Permeabilities of Common Ferromagnetic Materials (46)

Material	Relative Permeability(μ_r)
Cobalt	250
Nickel	600
Iron	6,000
Supermalloy(5%mo,79% Ni)	107
Steel (0.9 C)	100
Silicon iron (4% si)	7,000
Permalloy(78.5% Ni)	100,000
Fe3O4 (Magnetite)	100
Ferrites	5,000
Mumetal (75% Ni, 5% Cu, 2% Cr)	100,000
Permendur	5,000

Soft iron is used in electromagnetic applications like transformer cores, motors and temporary magnets. The next section is a summary on transformer basics and how electromagnetic energy is used to generate electricity.

3.3 An Overview of Electrical Transformers

This section looks at the basic working principle of transformers and the relevant equations for their design. A transformer is an electrical device that transfers energy from one circuit to another by electromagnetic induction (47). The key components of a transformer are a primary coil (winding), a secondary coil (winding) and a core material

that supports both windings. In addition, the core also acts as the medium for magnetic coupling between the primary and secondary windings. Thus, this can be said to be the medium for energy transfer. A simplified diagram of a transformer is shown in Figure 3.4.

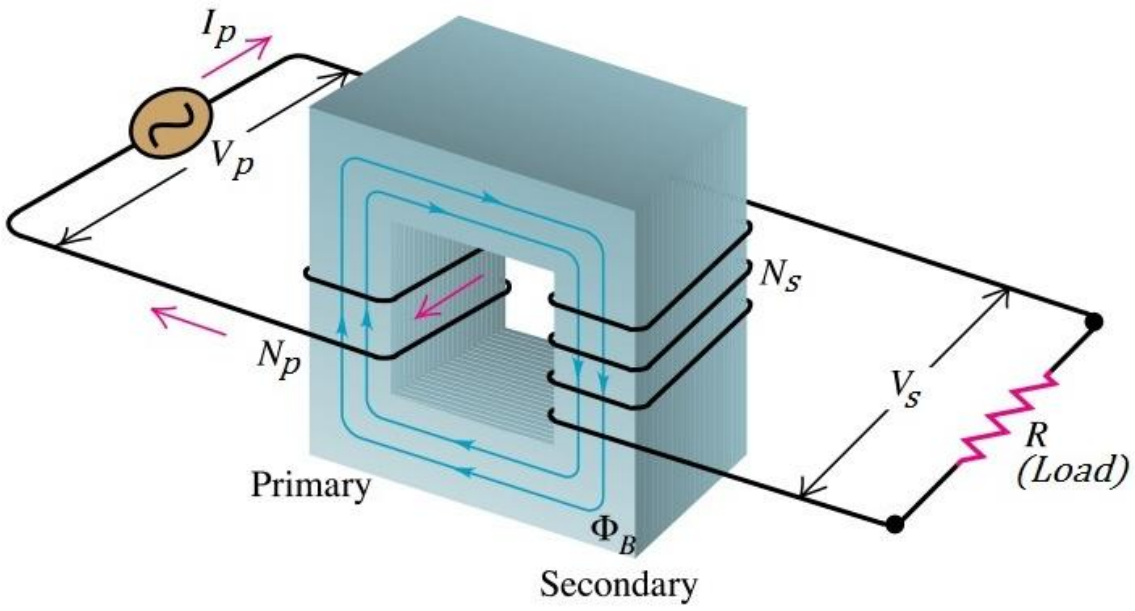


Figure 3.4: A Simplified Diagram of an Electric Transformer (5)

Faraday's law of induction is the basic law of electromagnetism relating to the operating principles of transformers. The law states that, time varying magnetic fields can produce an electric current in a closed loop (44). In other words, any coil of wire in a changing field of magnetic field will cause a voltage (electromotive force, emf) to be induced in the coil. The change can be a relative movement of either magnetic source or the coil. This is expressed in equation form as (44), (45):

$$\mathcal{E} = -N\frac{d\phi}{dt} = -N\frac{d}{dt}\int_s B \cdot ds \qquad 16$$

Where, $\mathcal{E}$ induced emf, B magnetic field strength, N number of coil turns, s surface area of the inductor and ϕ magnetic flux density.

Conventional transformers are either used to step up or down voltage from a primary side to the secondary side winding. A simple transformer circuit is shown in Figure 3.5. They are constant power devices which means, the product of the primary voltage and current is the same as the product of the secondary voltage and current. The ideal transformer equations are:

$$|Primary\ power| = I_p V_p = |Secondary\ power| = I_s V_s$$

$$\frac{V_s}{V_p} = \frac{N_s}{N_p} = \frac{I_p}{I_s} \qquad 17$$

V_p Primary side voltage, V_s Secondary Voltage, N_p Primary turns, N_s Secondary turns, I_p Primary current, I_s Secondary current.

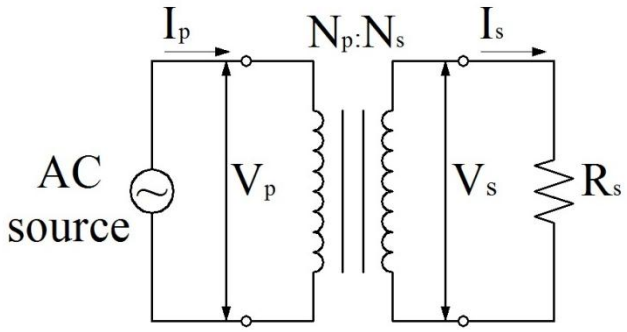


Figure 3.5: Simple Transformer Circuit

In practical transformers however, material properties of core plays an important role. The cores are selected so as not to saturate under low magnetic flux. Thus transformers are designed with cores of high permeabilities. The design principle is based on estimating the total component losses and performing an iterative process to get a balance between output performance and geometry of the core. The losses are copper

losses of the windings and losses of the core material (49). Practical transformers are designed by using equation 18 (49).

$$\frac{W_A(A_c)^{\left(\frac{2(\beta-1)}{\beta}\right)}}{(MLT)l_m^{(2/\beta)}}\left[\left(\frac{\beta}{2}\right)^{-\left(\frac{\beta}{\beta+2}\right)}+\left(\frac{\beta}{2}\right)^{\left(\frac{2}{\beta+2}\right)}\right]^{-\left(\frac{\beta+2}{\beta}\right)} = \frac{\rho\, I_{tot}^2\, \lambda_1^2\, K_{fe}^{\left(\frac{2}{\beta}\right)}}{4K_u(P_{tot})^{\left(\frac{\beta+2}{\beta}\right)}} \qquad 18$$

Where, ρ effective wire resistivity, I_{tot} total rms winding current, referred to primary, $\lambda_1$applied primary volt-sec, P_{tot} Allowed total power dissipation, K_u winding fill factor, β core loss exponent, K_{fe} core loss coefficient, A_c core cross-sectional area, W_A core window area, l_m magnetic path length, MLT mean length per turn of winding.

The terms on the left hand side of equation 18 are dependent on core geometry and that on the right hand side depend on the specifications of application (output voltage, current). Complete derivation of equation 18 and other related equations are found in (49) and (50).

3.3.1 Current Transformers (CT)

Harvesting electromagnetic flux around current carrying conductors through conventional voltage transformer design will require the current carrying conductors to be wound around transformer cores. This will involve modification of existing electrical wiring systems and also can be difficult to achieve with solid conductors. The current transformer (CT) is therefore the suitable choice to harvest energy from the magnetic flux around current carrying conductors. This section looks at the main theoretical equations used for designing current transformers.

Current transformers are used in high-power circuits where large currents are involved. They are mostly used in monitoring current of conductors or cables in AC power circuits (50). Figure 3.6 shows a circuit diagram of a simplified current transformer used to monitor line currents.

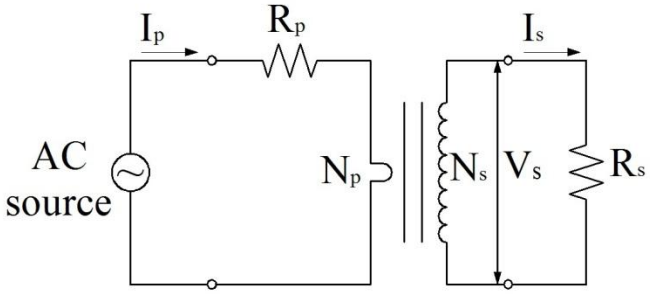


Figure 3.6: Simplified Circuit Diagram of a Current Transformer

Lists of predominant equations for designing current transformers are included below.

Application specific equations (50):

$\frac{I_p}{I_s} = \frac{N_s}{N_p}$, $N_p = 1$

$$\frac{I_p}{I_s} = N_s \quad (19)$$

$$I_s = \frac{V_s}{R_s} \text{ (A)} \quad (20)$$

I_p Primary current, I_s secondary current, N_p primary turns, N_s secondary turns, V_s secondary load voltage, R_s secondary load resistance.

For an output DC load R_o, at the secondary side, a bridge rectifier will be required to convert the secondary voltage from AC to DC voltage. Figure 3.7 shows the circuit diagram of a bridge rectifier incorporated into the secondary side of the CT.

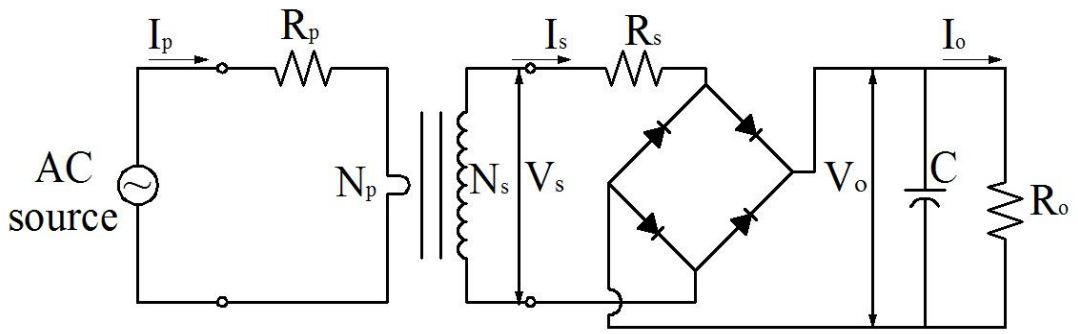


Figure 3.7: Complete Circuit Diagram of a Current Transformer with a Bridge Rectifier

Secondary winding resistance R_s

$$R_s = MLT\ (N_s)\left(\frac{\mu\Omega}{cm}\right)(10^{-6})\ (\Omega) \qquad 21$$

MLT is mean length per winding turn

and the output load voltage, V_o

$$V_o = I_o(R_o) \qquad 22$$

The Secondary voltage V_s is calculated using equation 23

$$V_s = V_o + V_d \text{ (Volts)} \qquad 23$$

Where, V_d is diode voltage drop

Total secondary output power, P_o

$$P_o = I_s V_s \text{ (watts)} \qquad 24$$

Core Geometry Equations (50):

Figure 3.8 illustrates current transformer core geometry parameters. The window area W_a is a through rectangular hole in the core material.

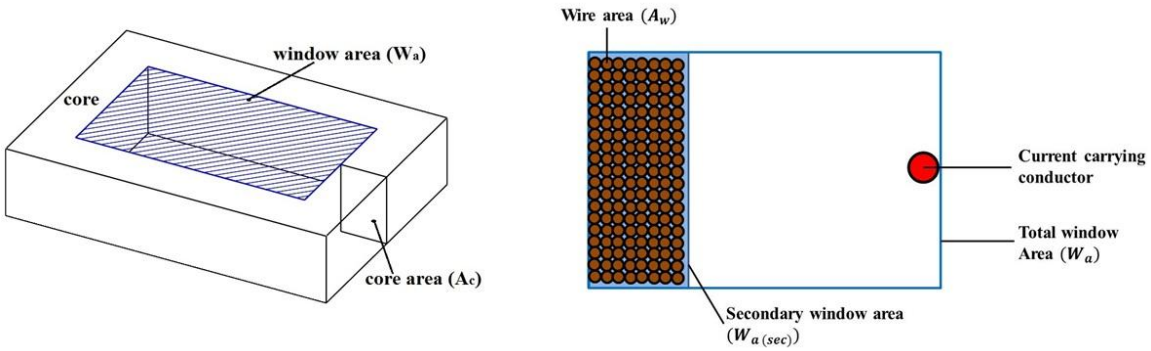


Figure 3.8: Current Transformer Core and Window Area

Core cross sectional area A_c

$$A_c = \frac{V_s(10^4)}{K_f B_{ac} f N_s} (\text{cm}^2) \quad 25$$

K_f is Waveform factor (4 for square wave and 4.44 for sinusoidal wave), (50). B_{ac} is Operating Magnetic flux density, f is Operating frequency,

Effective window area, $W_{a(eff)}$

$$W_{a(eff)} = W_a S_3 (\text{cm}^2) \quad 26$$

W_a is Core window area, S_3Effective window area constant (ratio of usable window area to window area), typical value of $S_3$is 0.873 (31).

Secondary window area, $W_{a(sec)}$

$$W_{a(sec)} = \frac{W_{a(eff)}}{2} (\text{cm}^2) \quad 27$$

Wire area, A_w

$$A_w = \frac{W_{a(sec)} S_2}{N_s} (\text{cm}^2) \quad 28$$

$S_2$is wire lay fill factor (ratio of wound area to usable window area), typical value of S_2 is 0.61 (50).

Table A.1, (50) in the appendix shows various wire areas (A_w) and corresponding AWG wire for winding. The general rule for wire selection is that when calculated, the wire area should not fall within 10% of the listed value in the wire table, otherwise the next smaller size should be selected. The design of other special CTs are found in (50).

3.3.2 Transformer Core Saturation

Transformer performance is limited by the amount of magnetic flux that can be contained in the core. Ferromagnetic cores tend to saturate after a certain level of magnetic flux. This further increase of the magnetic field flux after saturation will not result in a proportional increase in induced voltage. Saturation causes distortion of the wave shape of the primary current or voltage to secondary windings. This results in what is called harmonics in transformers. Harmonics are the distortions in wave shape of secondary side AC voltages. Common indication of harmonics in transformers is increase in temperature of during operation.

In order to avoid harmonics in transformers, material of cores should have high permeabilities. Saturation is also based on the core size, the number of coil turns, as well as the voltage and current in the windings. To improve the permeability of the cores, cut-cores are often used in most transformer applications.

The small size of the energy harvesting device requires a core material that will not saturate under the magnetic flux generated by the operating current. The typical operating current target range for this research is 1-20A. The magnetic core material

proposed for the energy harvesting device is Metglas allow 2605SA1. Metglas 2605SA1 has a high saturation flux density of 1.56 T (51) which is very high compared to the maximum magnetic flux for the target 1-20A. Simulations of magnetic flux generation from various currents are discussed in Chapter 5. The next chapter gives the objectives of this research and explains the methodology involved. The design process of the proposed energy harvesting device is illustrated in a case study.

Chapter 4

RESEARCH OBJECTIVE, METHODOLOGY, AND DESIGN

4.1 Introduction

Developing a miniature power supply that has a limited source of energy (electromagnetic field) requires in-depth design considerations. Apart from the size and source of energy being a limitation, the minimum required amount of power that can be generated also depends largely on these limitations. This minimum amount of power should be able to power the sensor to enable continual operation. Therefore, a careful design has to balance the aforementioned limitations whilst also meeting the required minimum output power. This chapter discusses the main objectives, methodology, and design considerations that must be taken into account to meet such a goal.

4.2 Research Objective and Methodology

The main objectives and methodology of the research can be summarized as follows:

[1] To develop and design a small electromagnetic energy harvesting device that will fit into the standard electrical node boxes in Figure 4.1 (52). The most common sizes of boxes used for wiring electrical nodes are 95mm by 50mm by 90mm and 100mm sq. by 38mm. Therefore, the size of the harvesting device shall meet the existing form factor of standard boxes for electrical nodes without any modification to the electrical wiring system.

Table 314.16(A) Metal Boxes

Box Trade Size			Minimum Volume		Maximum Number of Conductors* (arranged by AWG size)						
mm	in.		cm^3	$in.^3$	18	16	14	12	10	8	6
100 × 32	(4 × 1¼)	round/octagonal	205	12.5	8	7	6	5	5	5	2
100 × 38	(4 × 1½)	round/octagonal	254	15.5	10	8	7	6	6	5	3
100 × 54	(4 × 2⅛)	round/octagonal	353	21.5	14	12	10	9	8	7	4
100 × 32	(4 × 1¼)	square	295	18.0	12	10	9	8	7	6	3
100 × 38	(4 × 1½)	square	344	21.0	14	12	10	9	8	7	4
100 × 54	(4 × 2⅛)	square	497	30.3	20	17	15	13	12	10	6
120 × 32	(4 11/16 × 1¼)	square	418	25.5	17	14	12	11	10	8	5
120 × 38	(4 11/16 × 1½)	square	484	29.5	19	16	14	13	11	9	5
120 × 54	(4 11/16 × 2⅛)	square	689	42.0	28	24	21	18	16	14	8
75 × 50 × 38	(3 × 2 × 1½)	device	123	7.5	5	4	3	3	3	2	1
75 × 50 × 50	(3 × 2 × 2)	device	164	10.0	6	5	5	4	4	3	2
75 × 50 × 57	(3 × 2 × 2¼)	device	172	10.5	7	6	5	4	4	3	2
75 × 50 × 65	(3 × 2 × 2½)	device	205	12.5	8	7	6	5	5	4	2
75 × 50 × 70	(3 × 2 × 2¾)	device	230	14.0	9	8	7	6	5	4	2
75 × 50 × 90	(3 × 2 × 3½)	device	295	18.0	12	10	9	8	7	6	3
100 × 54 × 38	(4 × 2⅛ × 1½)	device	169	10.3	6	5	5	4	4	3	2
100 × 54 × 48	(4 × 2⅛ × 1⅞)	device	213	13.0	8	7	6	5	5	4	2
100 × 54 × 54	(4 × 2⅛ × 2⅛)	device	238	14.5	9	8	7	6	5	4	2
95 × 50 × 65	(3¾ × 2 × 2½)	masonry box/gang	230	14.0	9	8	7	6	5	4	2
95 × 50 × 90	(3¾ × 2 × 3½)	masonry box/gang	344	21.0	14	12	10	9	8	7	4
min. 44.5 depth	FS — single cover/gang (1¾)		221	13.5	9	7	6	6	5	4	2
min. 60.3 depth	FD — single cover/gang (2⅜)		295	18.0	12	10	9	8	7	6	3
min. 44.5 depth	FS — multiple cover/gang (1¾)		295	18.0	12	10	9	8	7	6	3
min. 60.3 depth	FD — multiple cover/gang (2⅜)		395	24.0	16	13	12	10	9	8	4

*Where no volume allowances are required by 314.16(B)(2) through (B)(5).

Figure 4.1: Standard Metal Boxes Used in Electrical Wiring Systems (Table 314.16 (A), NEC 2008)

Method :

- Develop a method for designing a small electric current transformer by balancing key parameters – the core size, primary energy source, and output performance. The primary source will be the magnetic field generated around a conductor carrying 1-20A in a 120V system. This will be mainly based on transformer design basics and theories.
- Propose and produce design drawings of a small current transformer for electromagnetic energy harvesting. The design drawing will be generated in CAD and shall be a shop working drawing standard.

- Coordinate with the electric transformer manufacturing company to manufacture the proposed design.

[2] Develop a method for estimating the electromagnetic field strength around the current carrying conductors and the equivalent amount of power that can be generated by the energy harvesting device using theoretical equations.

Method :

- Estimate the amount of magnetic field strength around the current carrying conductors. This will be based on the theory of electromagnetics and use MATLAB tools.
- Use theoretical equations for transformer design basics to estimate the equivalent amount of power that can be generated from the magnetic field above. This will be done with MATLAB tools.

[3] Build a prototype board made up of typical electrical nodes and run experiments with the proposed energy harvesting device. The board will have a 120V- 20A electrical distribution wiring system with an energy management sensor at the node. Typical loads such as light and plug loads will be used in the experiments.

Method:

- Design an electric distribution layout with a 120V- 20A rated conductor comprised of 5 nodes (2 lighting nodes with 2 switching nodes and 3 - 120V duplex nodes).

- Build the prototype board from the design layout. An energy management sensor will be attached to one of the nodes.
- The energy harvesting device will be mounted on the current carrying conductor at a node and hard wired to the energy management sensor. A re-chargeable battery shall be incorporated into the board for storing generated power during periods in which the sensors are idle.
- Two additional low powered devices (LED and USB power outlets) shall be mounted on the board for demonstration purposes and hard wired to the energy harvesting device.

[4] Investigate potential challenges and other useful applications. These challenges will be in relation to:

- Design limitations,
- Code violations and
- Practical implementation.

A list of other low power devices in the built environment will also be presented.

The next section explains the design process behind the creation of the energy harvesting device.

4.3 The Design Process

To illustrate design process, this section presents a case study to design the proposed energy harvesting device. The theoretical design approach is based on that of a CT. The various parameters for design and operation are given as follows:

- Sensor energy specifications (current consumption, battery voltage, average power consumption) are shown in Table 4.1.
- Energy harvesting device output voltage, 5V DC.
- Diode voltage drop is 2V DC.
- The core size specifications (height, Length, width, window size) are shown in Figure 4.2.
- Operating primary current range is 1-20A rms and a voltage of 120 V AC rms.
- Operating frequency, sine wave 60Hz.
- Operating current carrying conductor size is #12 AWG.

Table 4.1: Sensor Energy Demand Specifications

Sensor	Current Consumption			Battery Voltage	Average Power Consumption
	Transmission mode	Reception mode	Sleep mode		
XBEE Zigbee/802.15.4 Modules (15)	50mA	50mA	10 μA	2.8V-3.4V	155mW

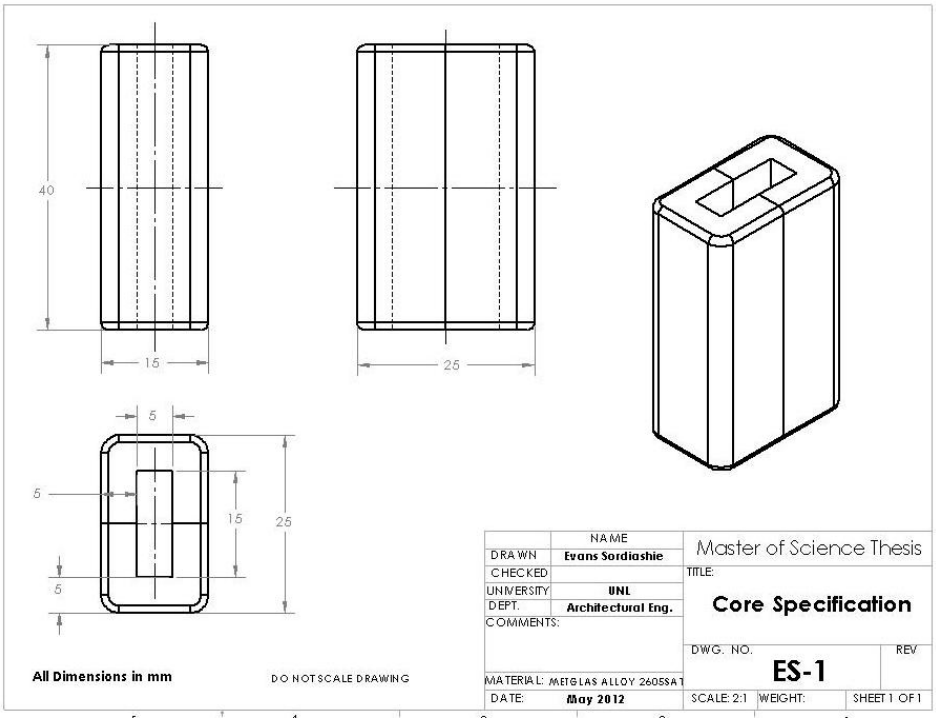


Figure 4.2: Energy Harvesting Device Core Specifications (proposed design)

4.3.1 Design Calculations

A cross sectional area of the core from equation 25 in Chapter 3 is given by;

$$A_c = \frac{V_s(10^4)}{K_f B_{ac} f N_s} \ (cm^2)$$

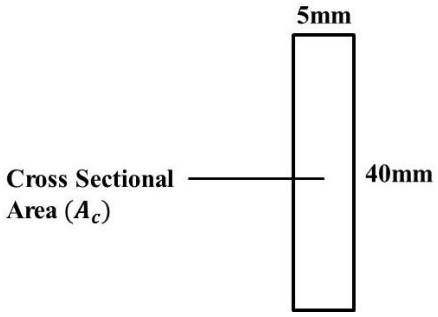


Figure 4.3: Cross-Sectional Area of the Core

From the core geometry in Figure 4.3,

Core height, $l = 40mm = 40 \times 10^{-3}\ m$

$A_c = (5 \times 40)mm^2 = 2\ cm^2$

Distance from current carrying conductor to magnetic path length r, as shown in Figure 4.4 is

$r = \left(\frac{5}{2} + 4\right) = 6.5\ mm = 6.5 \times 10^{-3}\ m$

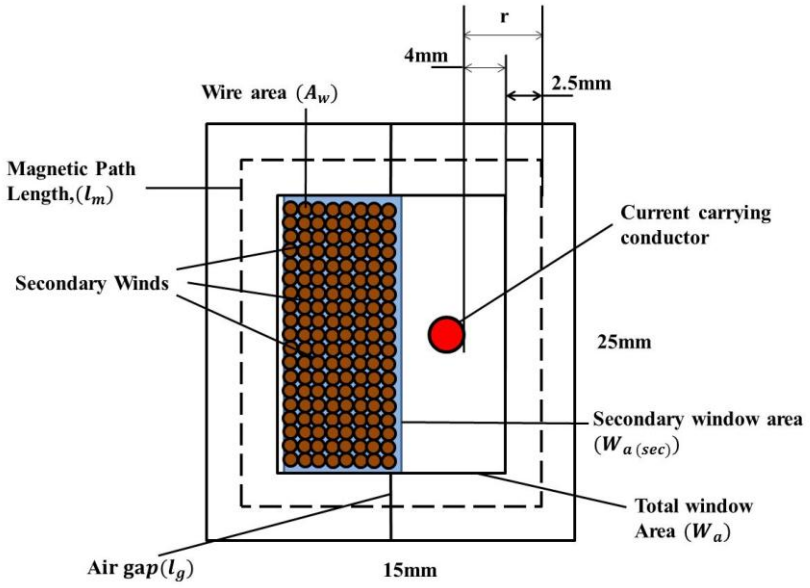


Figure 4.4: Core Window Area

Finding permeability of core

Permeability of free space $\mu_o = 4\pi\ \times 10^{-7}\ H/m$

Permeability of Metglas material, $\mu_m = 3000\ H/m$

Air gap between cut cores, $l_g = 120\ \times 10^{-6}\ m$

Magnetic path length, $l_m \cong 2\ (20 + 10) = 60 \times 10^{-3}\ m$

Overall relative permeability of Core assembly μ_r is given by (50);

$$\mu_r = \frac{\mu_m}{1 + \ \mu_m \left(\frac{l_g}{l_m}\right)}$$

$$\mu_r = \frac{3000}{1 + \ 3000 \left(\frac{120 \times 10^{-6}}{60 \times 10^{-3}}\right)} = 428.57$$

Finding magnetic field strength

Operating primary current, $I_p = 10\ A\ (rms)$

Magnetic field strength generated by *10 A* flowing through conductor is given by equation 14;

$$B_{ac} = \frac{\mu_0 \mu_r I_p l}{2\pi r \sqrt{4r^2 + l^2}}\ (T)$$

Where, $\mu = \mu_0 \mu_r$, effective permeability of core assembly.

$$B_{ac} = \frac{4\pi\ \times 10^{-7} \times 428.57 \times 10 \times (40 \times 10^{-3})}{2\pi \times 6.5 \times 10^{-3} \times \sqrt{4 \times (6.5 \times 10^{-3})^2 + (40 \times 10^{-3})^2}}\ (T)$$

$$B_{ac} = \ 0.1254\ T$$

Finding total secondary winding

Diode and secondary winding resistance voltage drop, $V_d \cong \ 3\ V$.

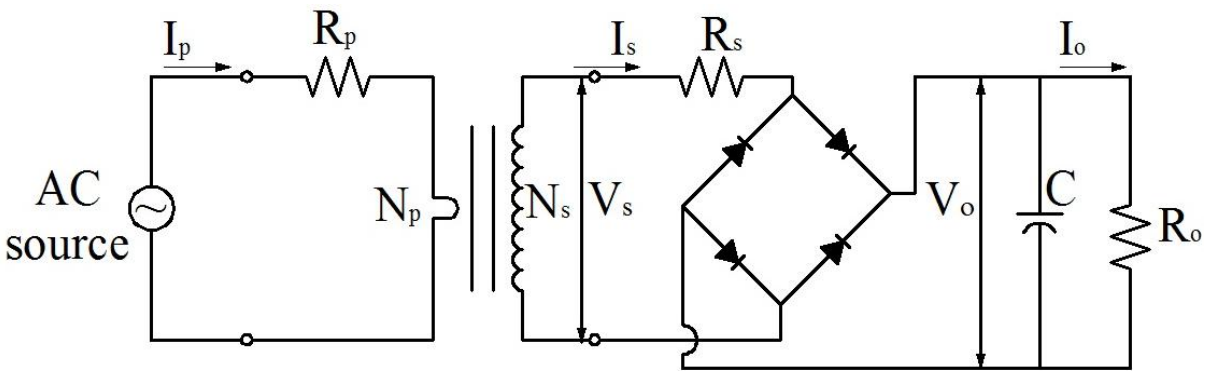


Figure 4.5: Circuit Diagram of the Energy Harvesting Device.

Expected output voltage is $V_o = 5\ V\ DC$

Output load current is$50\ mA$,

Corresponding output load resistance R_o is,

$R_o = \frac{V_L}{I_o} = \frac{3.4}{50} = 68\Omega$, V_L is actual sensor voltage requirement.

Total secondary voltage, V_s is given by;

$V_s = V_o + V_d$

$V_s = 5 + 3 = 8\ V(rms)$

Primary turns, $N_p \cong 1$

Waveform factor for sine wave, K_f = 4.44

Operating frequency, $f = 60\ Hz$

Total Secondary winding N_s, is given by;

$$N_s = \frac{V_s(10^4)}{K_f B_{ac}\, f A_c} = \frac{8 \times 10^4}{4.44 \times 0.1254 \times 60 \times 2} = 1197.3$$

$$N_s \cong 1200\ turns$$

Finding conductor size for winding

Total window area as shown in Figure 4.4 is,

$$W_a = 15 \times 5 = 75\ mm^2 = 0.75\ cm^2$$

Effective window area $W_{a(eff)}$, available for winding (both primary and secondary) is,

$W_{a(eff)} = W_a \times S_3$, S_3 is effective window area constant. Typical value of S_3 is 0.873 (50).

$$W_{a(eff)} = 0.75 \times 0.873 = .655\ cm^2$$

Secondary window area, $W_{a(sec)}$ is,

$$W_{a(sec)} = \frac{W_{a(eff)}}{2}\ (cm^2)$$

$$W_{a(sec)} = \frac{0.655}{2} = 0.327\ cm^2$$

Cross sectional area A_w of the conductor for winding is given by;

$A_w = \frac{W_{a(sec)} \times S_2}{N_s}$, S_2 is the fill factor or the wire lay, for usable window area. Typical value of S_3 is 0.61 (50).

$$A_w = \frac{0.327 \times 0.61}{1200} = 0.000166\ cm^2$$

Table A.1 (50) in appendix A shows various wire areas (A_w) and corresponding AWG wire for winding. The corresponding AWG conductor for the winding is selected to be # 36.

The corresponding winding resistance is calculated using equation 21.

$$R_s = MLT\ (N_s)\left(\frac{\mu\Omega}{cm}\right)(10^{-6})$$

MLT is the average perimeter of the cross sectional area of core before and after windings has been added. This is estimated as:

$$MLT = \frac{(40+5)\times 2 + (40+5)\times 2 \times 1.4 - 0.355}{2} \quad (mm)$$

Where, 140% is allowed for increase in perimeter for windings and 0.355mm is allowed for wrapper thickness on windings.

$$MLT = 10.8 \quad (cm)$$

$$R_s = 10.8\ (1200)(13608)(10^{-6}) = 176.36\Omega$$

The winding resistance is very high compared to the load resistance (68 Ω). To reduce this resistance, 500 turns of #25 AWG winding will be used for the secondary winding. This will also increase the current output of the harvesting device because of the constant power principle of the transformer. The new induced voltage V_s, is calculated using equation 25.

$$V_s = \frac{N_s \times K_f \times B_{ac} \times f \times A_c}{(10^4)} = 3.34\ V\ (rms)$$

The new winding resistance is,

$R_s = 10.8\,(500)(1062)(10^{-6}) = 5.73\Omega$

An AC to DC voltage doubler (multiplier) is used to boost up the new voltage. Figure 4.6 shows a circuit of the voltage doubler.

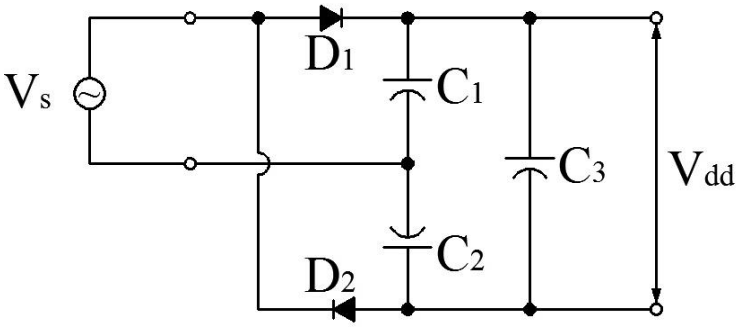


Figure 4.6: Voltage Doubler Circuit

The doubled voltage V_{dd} is calculated with equation 29

$$V_{dd} \cong 2 \times V_s\, peak = 2\sqrt{2} \times V_s\,(rms) \qquad 29$$

$$V_{dd} \cong 2\sqrt{2} \times 3.34 = 9.45\,V\,(DC)$$

To reduce and stabilize V_{dd} voltage for sensor operation voltage (2.8-3.4V DC) a voltage regulator is incorporated into the circuit. This is shown in Figure 4.7.

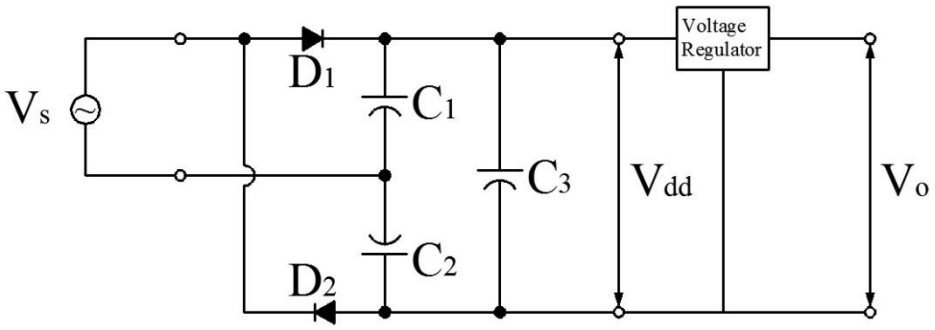


Figure 4.7: Voltage Regulator Incorporated into the Voltage Doubler Circuit

A summary of the final design specifications are as follows:

- Operating current carrying conductor, #12 AWG

- Primary turn, 1
- Average operating primary current , 10 A rms
- Operating frequency, sine wave 60Hz.
- Operating flux density, 0.1254 T
- Harvesting device size and specifications; Magnetic Material (Metglas alloy 2605SA1, High Saturation Flux Density (1.56 T)), Window area,W_a, 0.75 cm^2, Magnetic path length, $l_m \cong 6$cm, (see Figure 4.2).
- Winding conductor, # 25 AWG, 500 turns.
- Output voltage for sensor, 3.3V DC
- Output Load Resistance, ~ 68Ω
- Voltage regulator, LM2937 rated at 3.3 V DC
- Diode D_1, D_2 (IN4003), C_1 C_2 C_3 ,(1μ F rated at 100V)

This concludes the design process. The core size in Figure 4.2 is proposed core geometry for this research. The geometry will fit into typical node boxes (95mm by 50mm by 90mm or 100mm sq. by 38mm) in the built environment. In addition, management sensors can be directly mounted on the harvesting device at the node.

However, due to long manufacturing lead time for the proposed core of energy harvesting device, a smaller standard core readily available is used for prototyping and experiment. The core used is Amorphous Metal C Core #4 (AMCC-4). Detail specifications of the AMCC-4 core are provided in chapter 5.

Chapter 5

RESULTS AND DISCUSSION

5.1 Introduction

The main objective of this research is to design an electromagnetic harvesting device that will draw energy from available electromagnetic field around current carrying conductors at electrical nodes to generate power for energy management sensors. The size of this harvesting device is also a key objective as the device should fit into boxes at the electrical nodes without affecting the form factor of the boxes. To achieve these goals, it is very important to estimate the electromagnetic energy around typical current conductors in the built environment. With estimates of magnetic fields available, further estimates of equivalent voltage to be induced can be successfully done in relation to the size of the energy harvesting device core. The cross sectional area, A_c of the core needs to be carefully selected to prevent magnetic flux saturation but yet fit into typical boxes of nodes. This chapter presents simulation results and discusses the available electromagnetic energy around typical current carrying conductors in the electrical distribution system. All simulations are done with MATLAB software tools. Refer to Appendix C for the related programming codes.

The proposed core size in Figure 4.2 is not a standard core size available on the market. This means that a core must be customized for the proposed energy harvesting device. The manufacturing lead time to produce such a core is estimated to be 3 to 5 months (53). In addition, since only a few cores are needed for experiments and testing,

the initial cost will be high. To resolve this problem, a standard small size core is used in the simulations and experiments. Appendix B gives the winding specifications for the AMCC-4 core. All the results in this thesis are therefore based on the AMCC-4 core. This chapter also includes results from actual experiments conducted with an AMCC-4 core prototype energy harvesting device.

5.2 Electromagnetic Energy Around Current Carrying Conductors

Figures 5.1 and 5.2 show simulations of magnetic flux density around a current carrying conductor. Magnetic flux density reduces radially with increasing distances away from the conductor. This validates the Biot – Savart law which basically gives magnetic flux density around the conductor to be directly proportional to the current flowing through the conductor and inversely proportional to the distance of that point from the conductor.

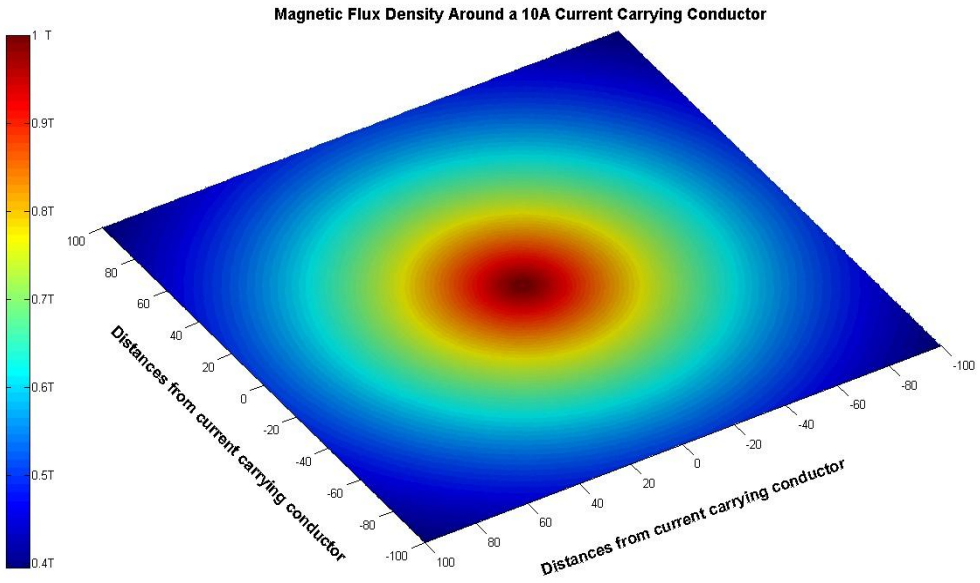


Figure 5.1: Magnetic Flux Density Around a 10 A Current Carrying Conductor

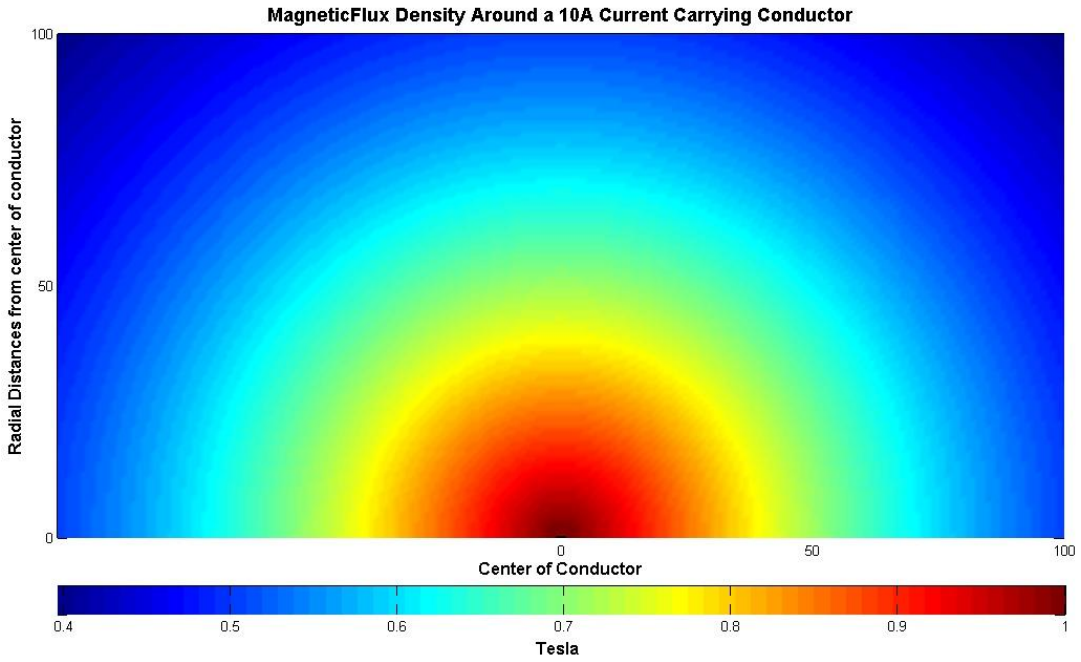


Figure 5.2: A Cut Through Section of Magnetic Flux Density Around a 10 A Current Carrying Conductor

Figure 5.3 shows theoretical result of electromagnetic flux density around a common current carrying conductor in the built environment. The conductor is #12 AWG and the current carrying capacity is 1-20A.

The magnetic flux density on the graph (see Figure 5.3) is calculated using permeability of free space ($\mu_o = 4\pi \times 10^{-7}$ H/m) and equation 14. To estimate magnetic flux that can be harvested for power generation, the relative permeability (μ_r) of the core has to be calculated as was done in the case study in chapter 4. Magnetic flux density around current carrying conductor decreases with increasing radial distances from the conductor. The simulation on the graph (see Figure 5.3) shows a radial distance of 2mm to 30mm from the current carrying conductor.

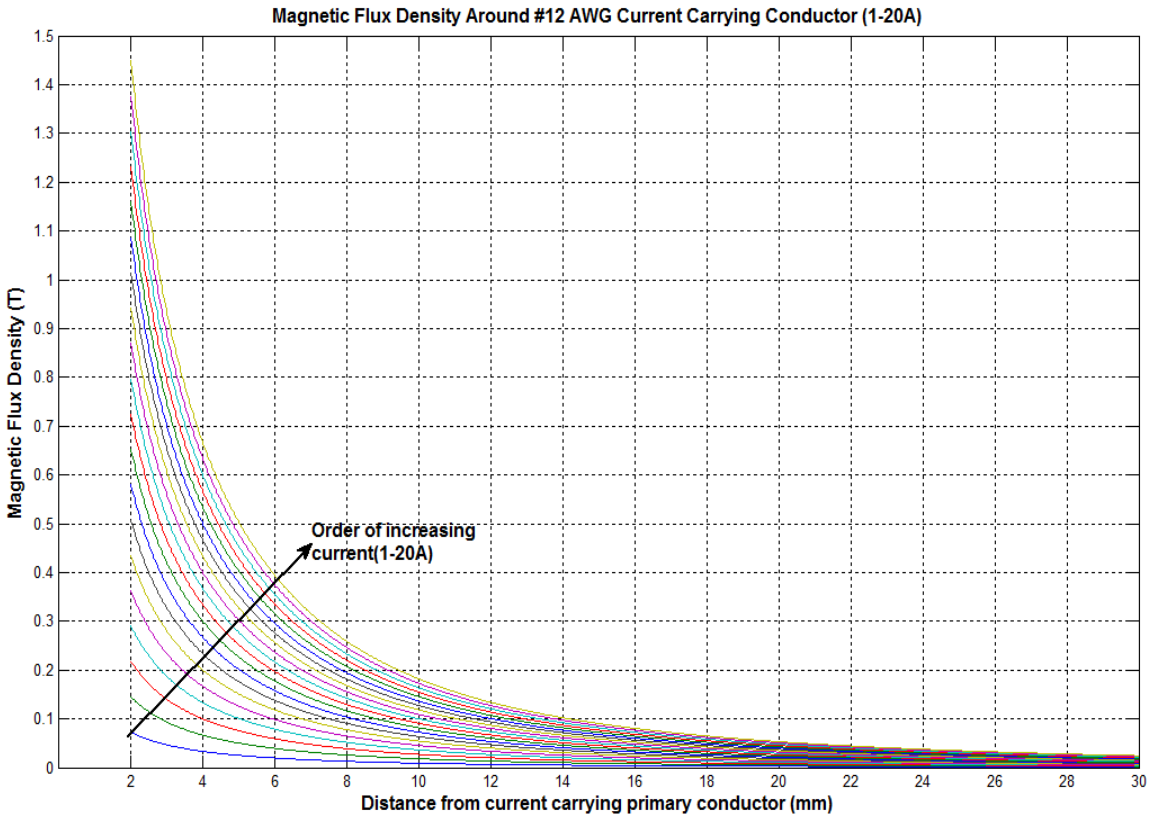


Figure 5.3: Magnetic Flux Density for Different Currents from 1-20 A

5.3 Induced and DC Output Voltages

The electric voltages that can be generated from corresponding primary load input (in watts) are shown in Figure 5.4. The voltages are generated using the AMCC-4 core. Each graph shows 5 different plots. The plots correspond to the voltages generated at radial distances from current carrying conductor to the magnetic path line (MPL, see Figure 4.4). The MPL is the length of an average magnetic field line around the interior of the core (49). The location of this field line is approximated to be at the center of the core as shown in Figure 4.4 (49). Figure 5.5 shows corresponding voltages for primary input current.

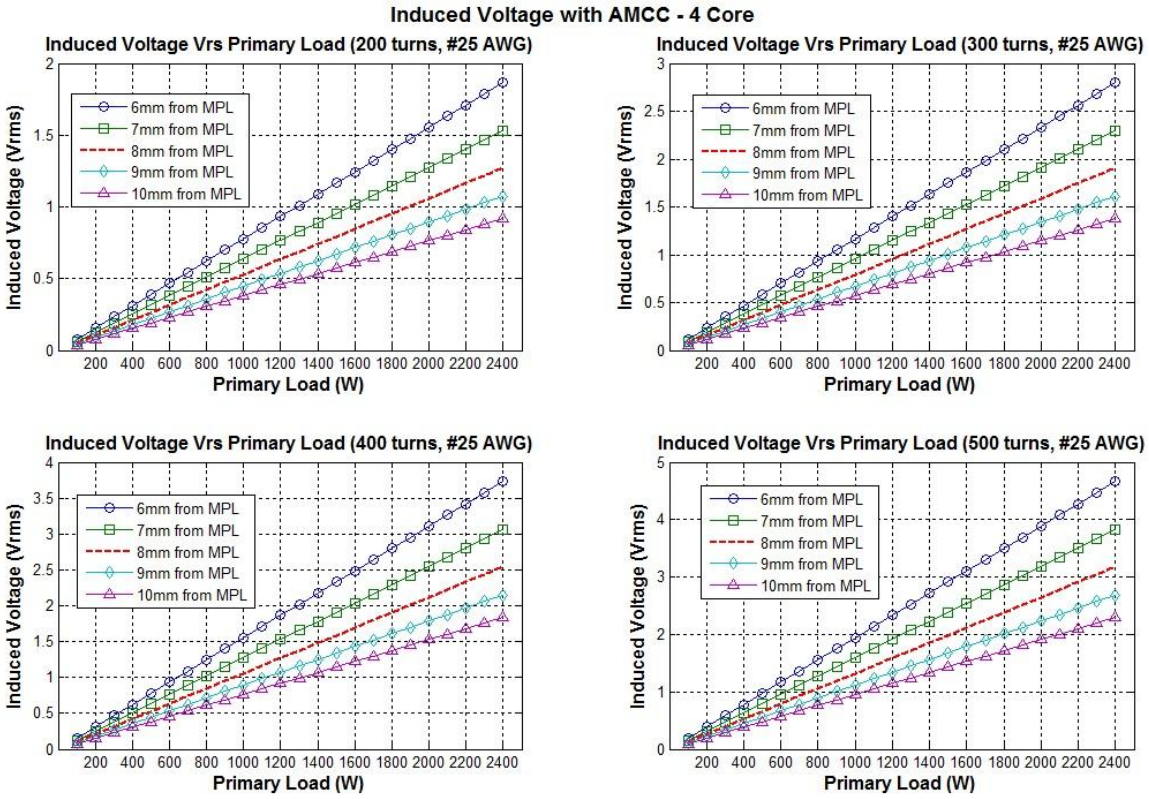


Figure 5.4: Induced Voltages per Primary Load Input (Watts)

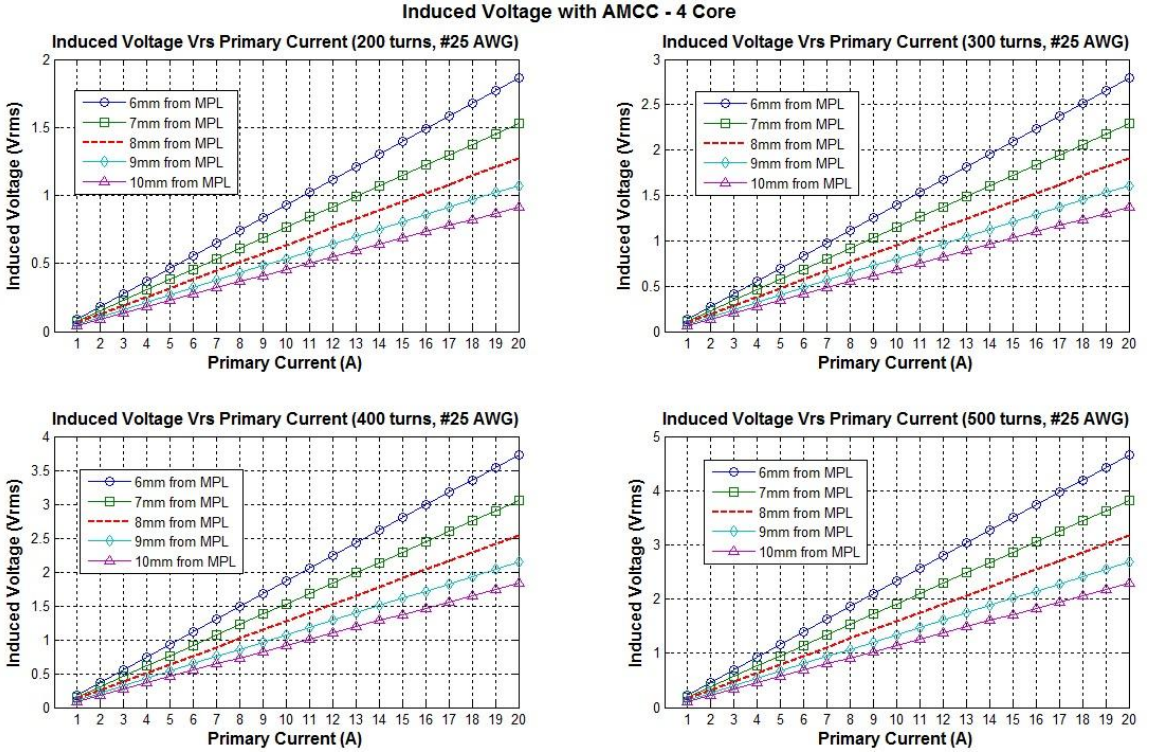


Figure 5.5: Induced Voltages per Primary Load Currents (A)

Figure 5.6 and 5.7 shows the voltages induced for a given load (watts) or current (A) and radial distances from current carrying conductor to MPL. The windings used on the AMCC-4 core for these simulations are 300 and 500 turns respectively.

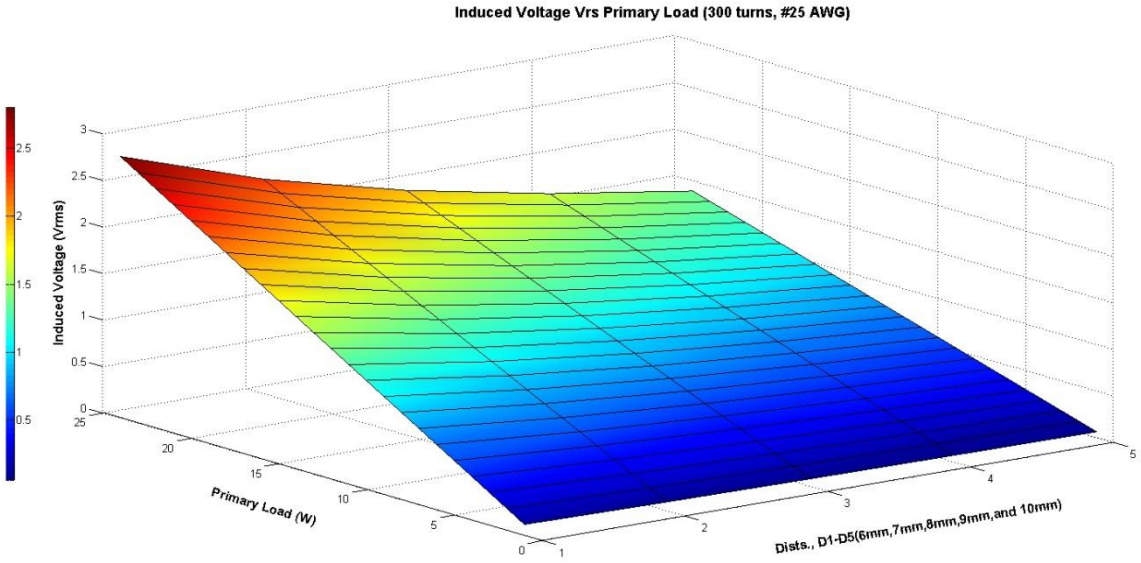


Figure 5.6: A Combination Plot of Induced Voltage Against Primary Input Load and Radial Distances from MPL (300 turns of #25 Secondary Winding)

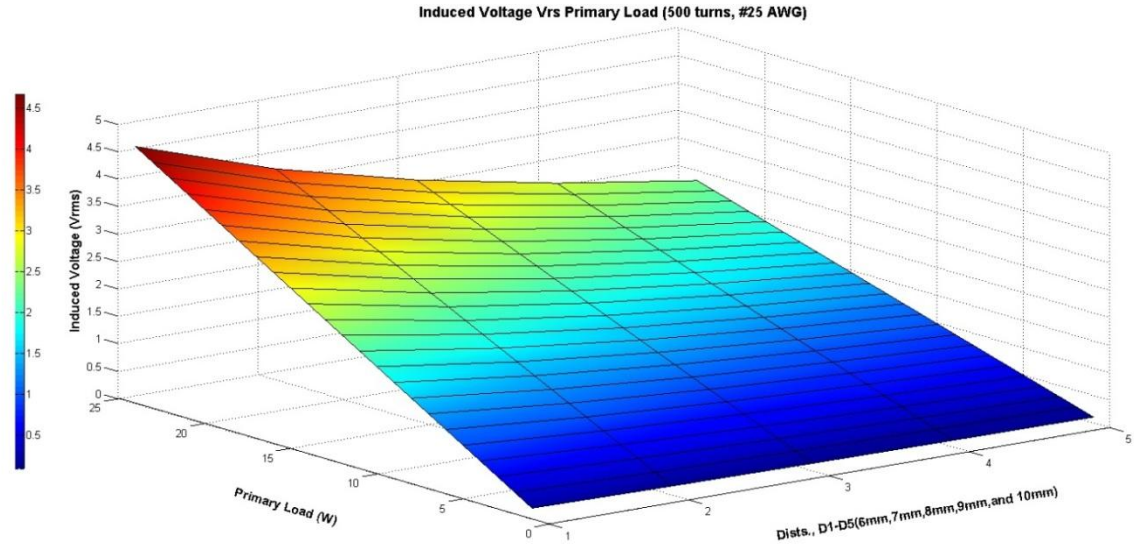


Figure 5.7: A Combination Plot of Induced Voltage Against Primary Input Load and Radial Distances from MPL (500 Turns of #25 Secondary Winding)

Most energy sensors use DC voltages during operation. Simulated voltages in Figure 5.5 are converted to DC component by incorporating an AC to DC bridge rectifier. A 0.5V voltage drop is allowed for the conversion from AC to DC. Figure 5.8 shows the DC voltage simulations.

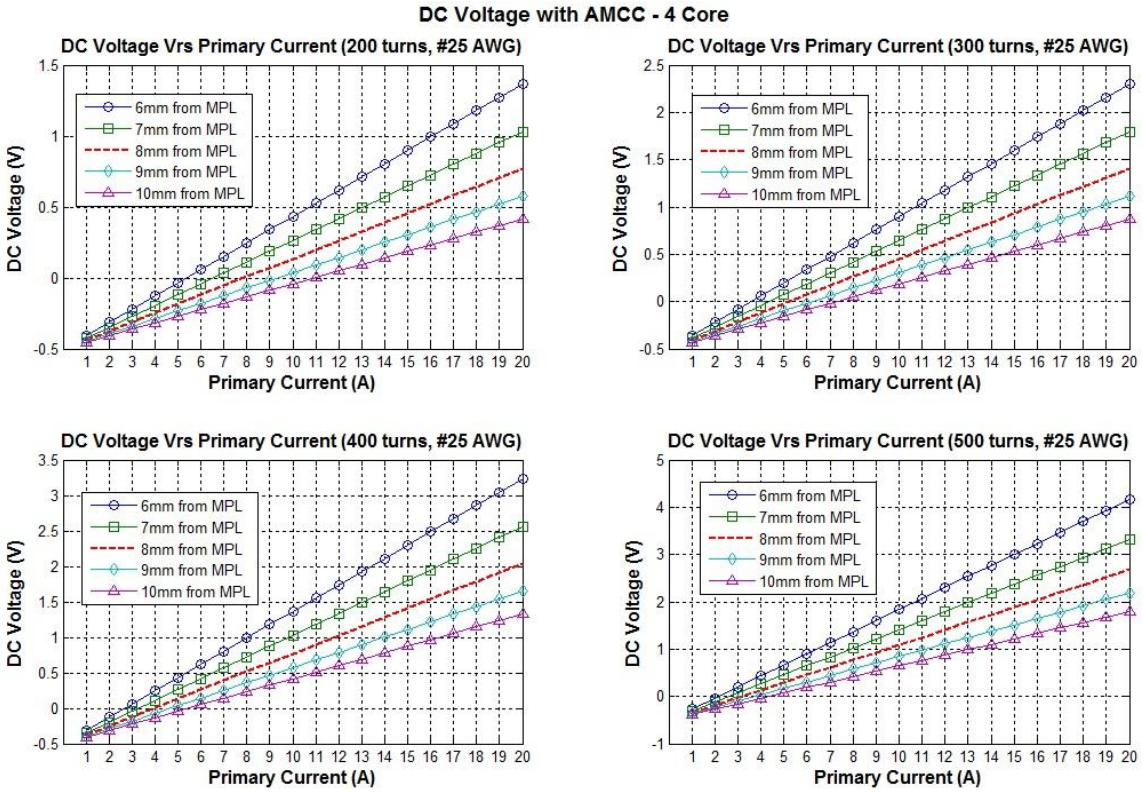


Figure 5.8: DC Output Voltages per Primary Input Load Currents

5.4 Core Size, Versus Magnetic Flux and Input Load Current

As one of the key objectives of the research, estimating the size of the core of energy harvesting device was very important. The size of the core basically translates to the cross sectional area of the core. Figure 5.9 shows a simulation of cross section estimation for a given magnetic flux density. The expected output voltages with a specific number of secondary windings on the core are also given. Corresponding graph for core cross section and primary input current is also shown in Figure 5.10.

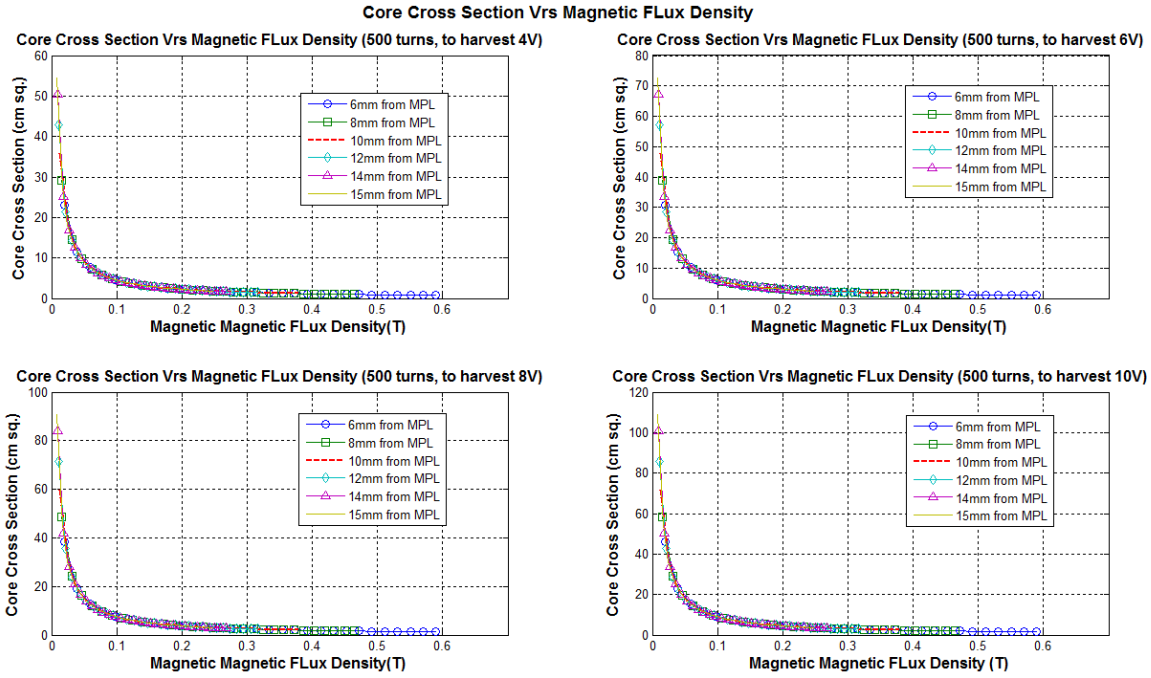


Figure 5.9: Core Cross Section vs. Available Magnetic Flux Density

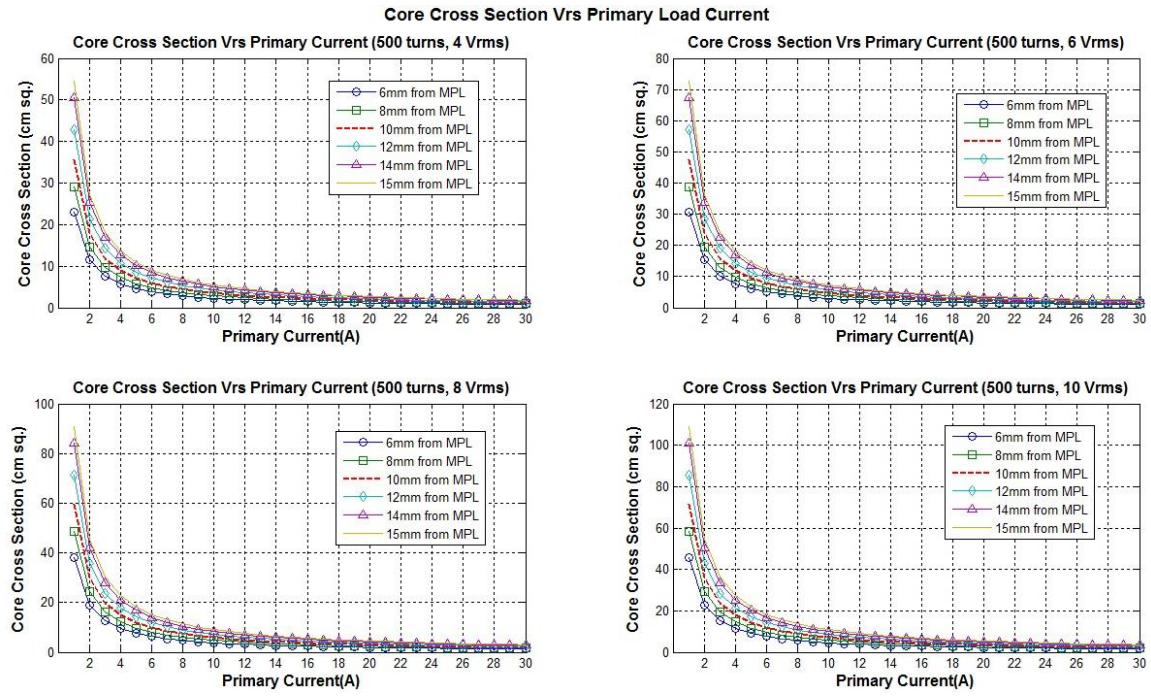


Figure 5.10: Core Cross Section Vs. Primary Input Current

To analyze how radial distances and input primary current affect the cross sectional area of the core of harvesting device, a combination plot is presented with radial distances and primary currents as inputs. This is shown in Figure 5.11.

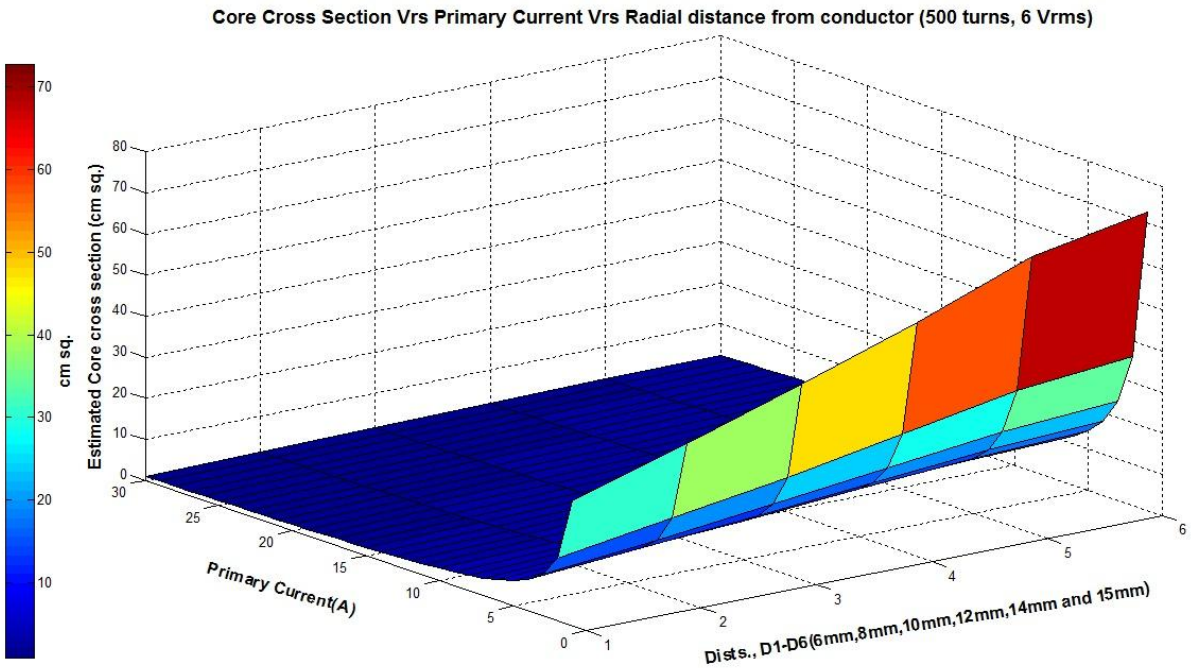


Figure 5.11: Combination Plot of Cross Sectional Area vs. Primary and Radial Distances

5.5 Experimental Results

To validate results from simulations, a prototype energy harvesting device is built with an AMCC-4 with windings, 200, 300, 400 and 500. Figure 5.12 shows an image of the AMCC-4 with the leads of the various windings. Except for the black colored, each lead corresponds to a number of windings. A circuit diagram of the windings is presented in Appendix B2.

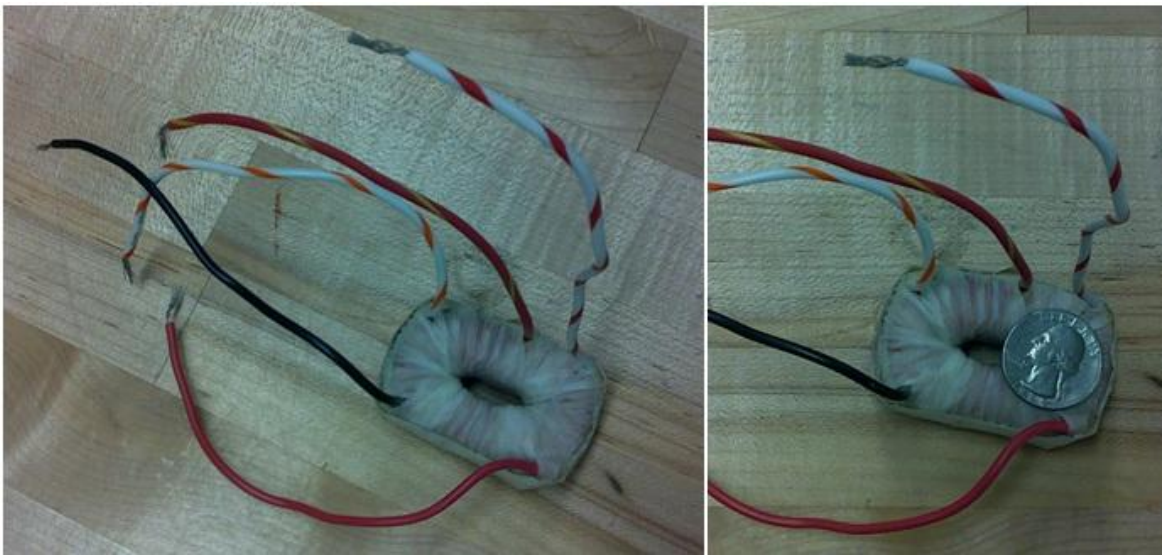


Figure 5.12 Prototype Energy Harvesting device made from AMCC-4 Core.

The board used for experiments is in Figure 5.13. The board is made up of some typical nodes in the built environments and rated for 120 V. The harvesting device is mounted on

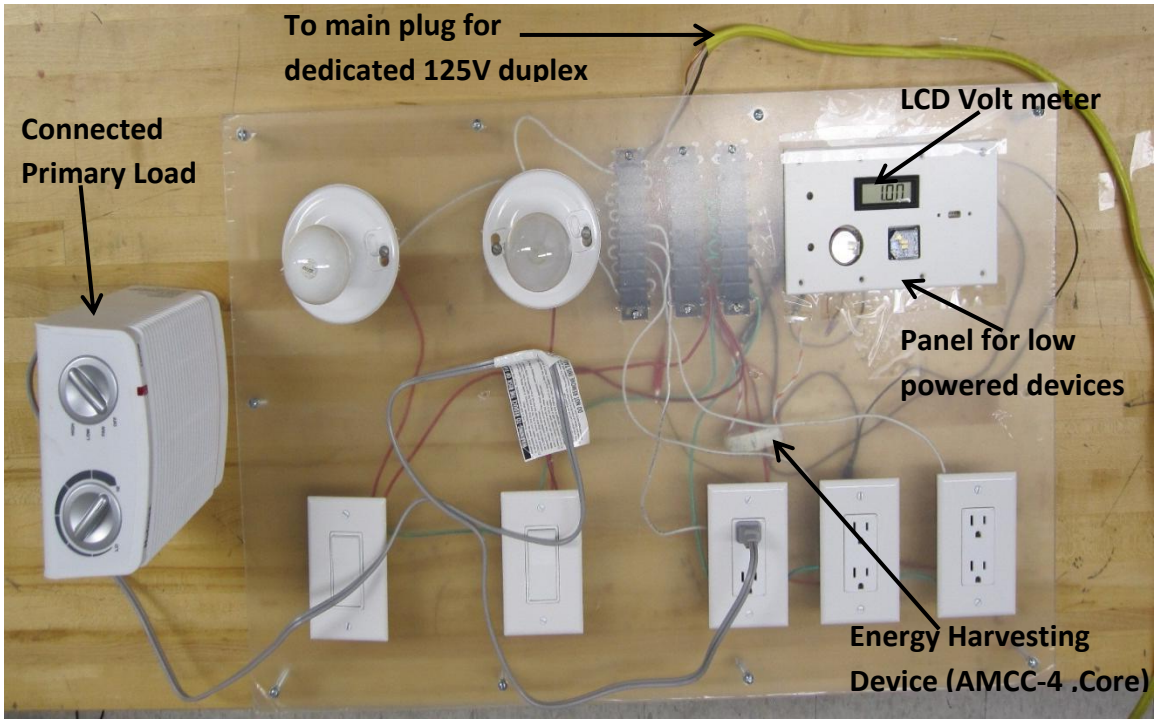


Figure 5.13: Board for Energy Harvesting Experiments

a current conductor on the board as shown in Figure 5.13. The board consists of duplex receptacles, lamp sockets and switches and low powered devices. A main plug for dedicated duplex receptacle is also included to supply power to the board.

The following are part inventory used in building the board for experiments.

1) Plastic platform
2) 3 duplex receptacles rated at 125 V.
3) 2 lamp holders rated at 120 V with connecting switches.
4) All conductors are #12 AWG (single phase, Neutral and Ground)
5) Low powered devices (energy sensors, USB outlet, LED lights, LED light)
6) DC bridge voltage multiplier (Diode D1, D2 (IN4003), C1 C2 C3 , 1μ F rated at 100V)
7) Voltage regulator, LM2937 rated at 3.3 V DC, LCD voltage display meter
8) 3V DC Chargeable Li-ion battery
9) #12 AWG main electric plug connector for a dedicated duplex receptacle rated at 125V.

Figure 5.14 shows a schematic diagram of the board used for experiment. The following steps are used to take measurements from the board.

a. The main plug is plugged into a dedicated duplex receptacle rated at 120/125V.
b. A multimeter is connected to the leads of the harvesting device installed on the current carrying conductor on the board.
c. Various sizes of loads (lamps, small size heaters, etc.) are connected to the nodes (lamp socket, receptacles) on the board.

d. The loads are turn on and induced voltage measurements are recorded from the multimeter. The sizes of the loads are also recorded. This step is repeated for various loads up to a combined total of 2400 W and also for different winding turns of the harvesting device.

e. The recorded induced voltage measurements are plotted against corresponding connected loads for the different windings on the harvesting device.

f. The main plug is disconnected from the dedicated duplex receptacle to conclude the experimental process.

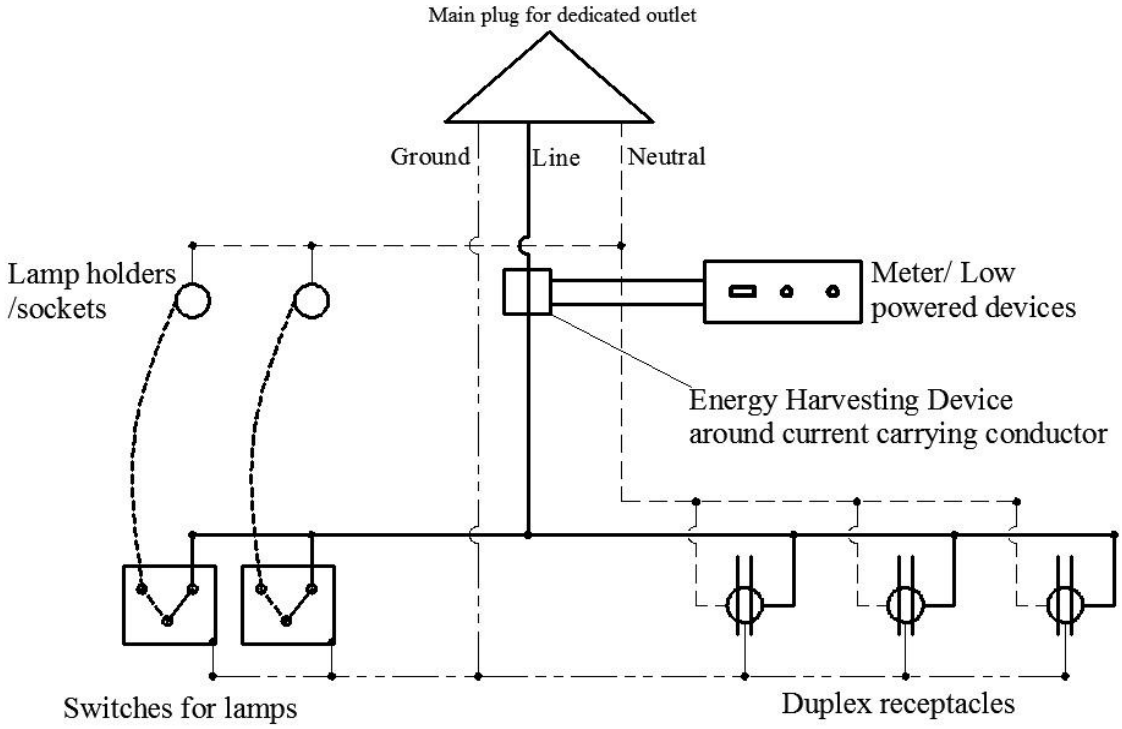


Figure 5.14: Schematic Diagram of the Experimental Board

Figure 5.15 and 5.16 show plotted measurements from experiments with different winding configurations for the AMCC-4 harvesting device.

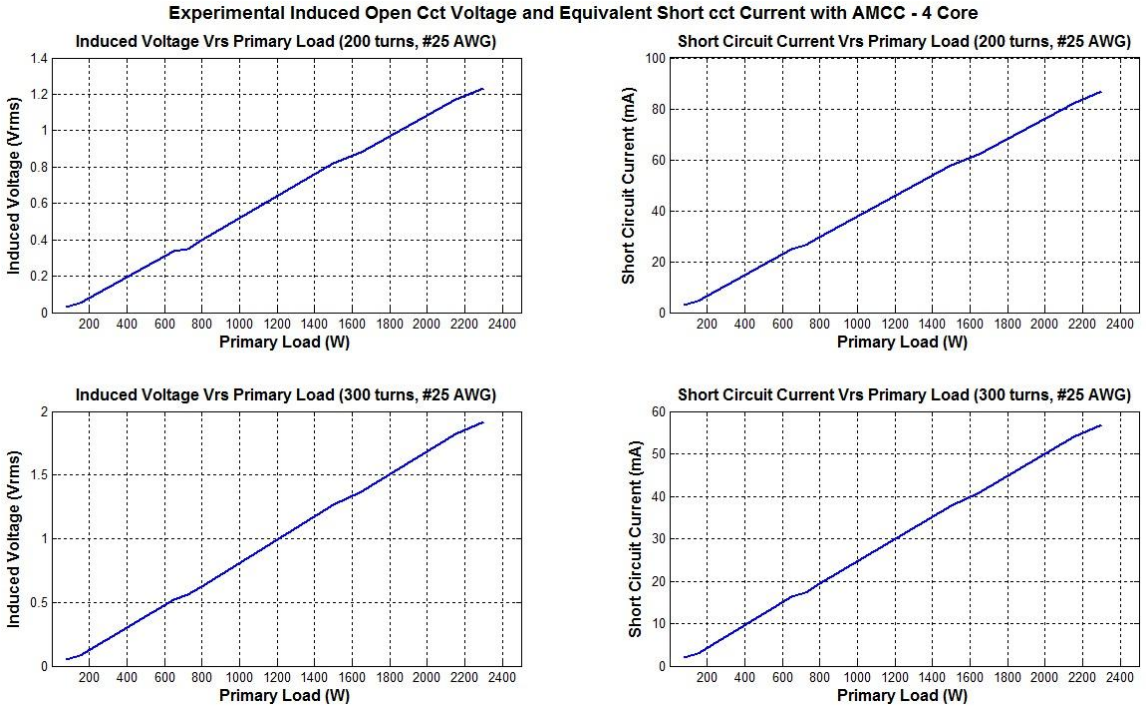


Figure 5.15: Experimental Induced Voltages with AMCC-4 Core

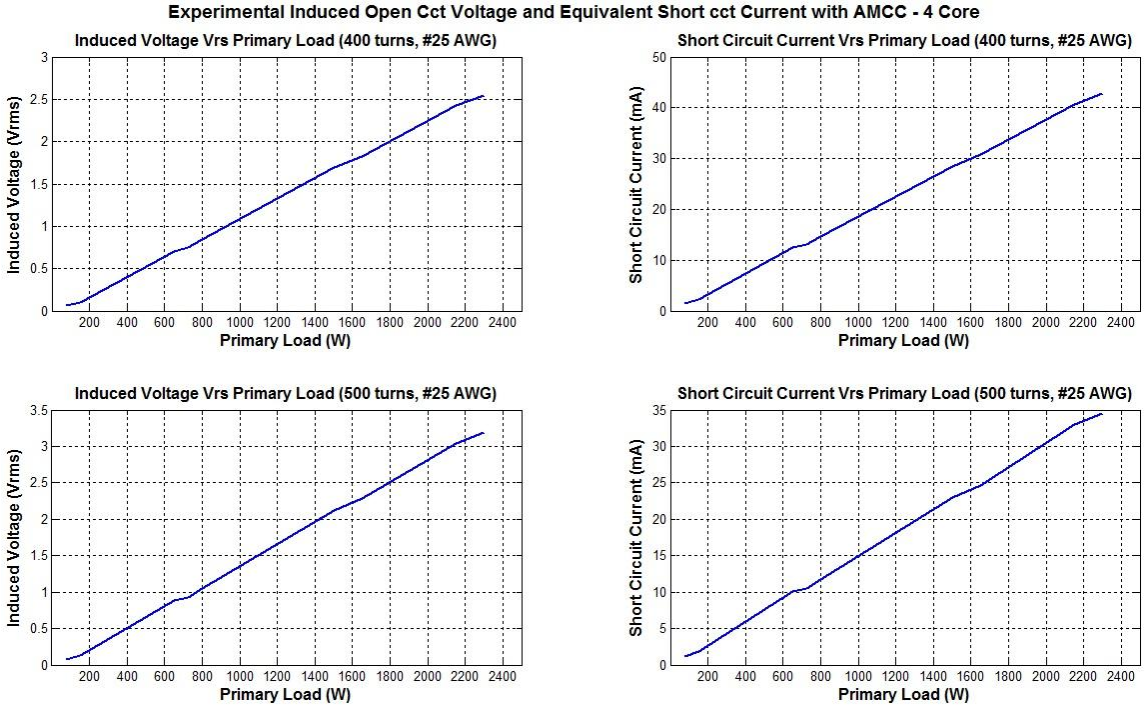


Figure 5.16: Experimental Induced Voltages with AMCC-4

5.6 Simulation vs. Experimental Results

A comparison between simulated and experimental induced voltages for various turns of secondary windings is shown in Figure 5.17. Experimental measurement are plotted with a black line and labeled on each graph.

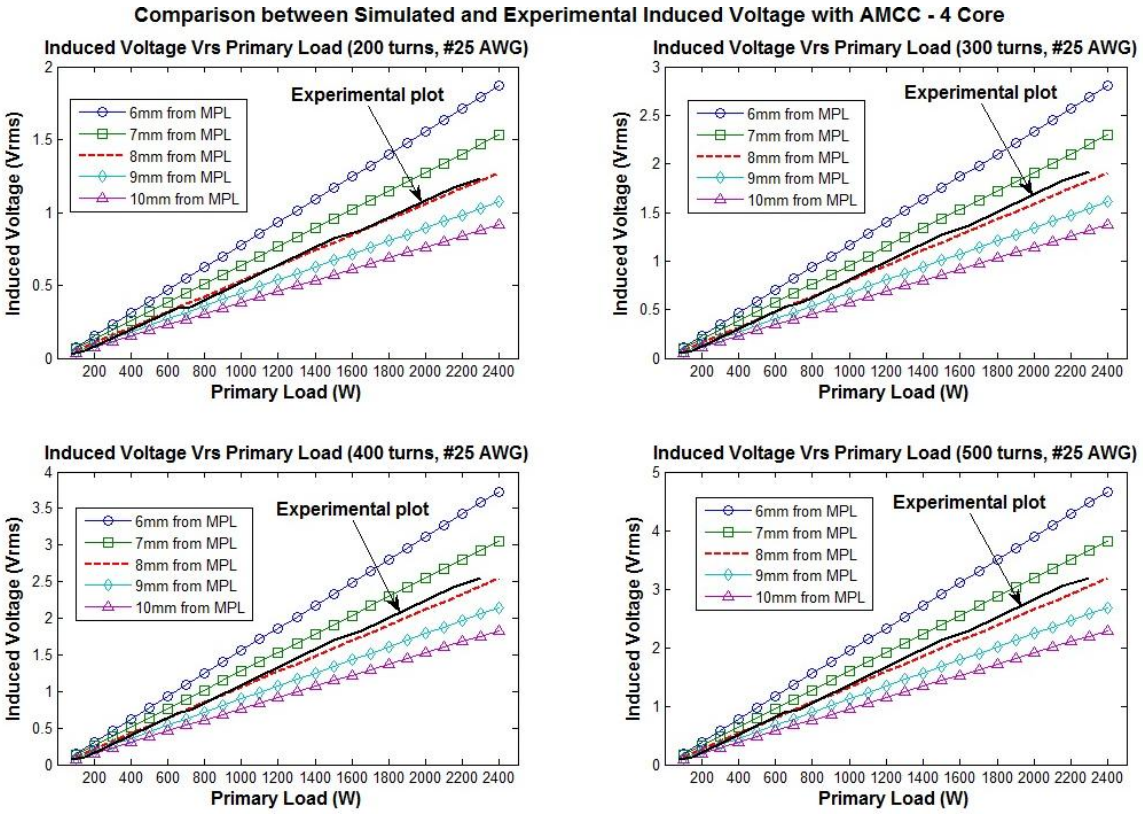


Figure 5.17: Comparison of Simulated and Experimental Induced Voltages

Table 5.1: Simulated and Experimentally Measured Induced Voltages (r = 8mm)

Primary Load (Watts)	Induced Voltages (V rms)							
	200 Turns		300 Turns		400 Turns		500 Turns	
	Simulated	Measured	Simulated	Measured	Simulated	Measured	Simulated	Measured
150	0.08	0.05	0.12	0.08	0.16	0.10	0.20	0.13
650	0.34	0.34	0.52	0.52	0.69	0.70	0.86	0.89
800	0.42	0.40	0.64	0.62	0.85	0.84	1.06	1.05
1500	0.80	0.82	1.19	1.27	1.59	1.70	1.99	2.12
1650	0.87	0.88	1.31	1.37	1.75	1.83	2.19	2.28
2150	1.14	1.17	1.71	1.82	2.28	2.43	2.85	3.04
2300	1.22	1.23	1.83	1.92	2.44	2.55	3.05	3.19

Table 5.1 shows the values of simulated and measured induced voltages from experiments. The voltages are tabulated according to the sizes of loads drawing current at the electrical nodes. Simulated and measured voltages correspond to an 8mm radial distance from the primary current carry conductor to the MPL. The measurements and simulations are also tabulated according to the number of winding turns on the AMCC-4 core.

5.7 Discussion

The objective of this research is to design an energy harvesting device to scavenge electromagnetic energy in the built environment for the purposes of supplying power to energy management sensors. The average power requirement of energy management sensors is 2V – 6 VDC and 12 – 170 mA DC. The amount of power generated depends on the proximity of the current carrying conductor to the MPL. A comparison between the simulation and experimental results indicate that the current carrying conductor is approximately 8mm in proximity from the MPL. With this proximity, simulation and experimental results show a little over 3VAC can be generated with an AMCC-4 core with 500 turns of #25 AWG. An equivalent short circuit current of 34mA AC was measured in laboratory experiments. Using a bridge voltage doubler or multiplier, the induced 3VAC can be boosted up to 6VDC to be feed into a voltage regulator and stabilized for sensor input.

The AMCC-4 was used in the simulation and experiments since it was the smallest available cut core on the market with the preferred magnetic permeability for this research.

The AMCC-4 core has a cross section of 1.11 cm^2 which is much smaller than that of the proposed core in the case study in Chapter 4. The height of the AMCC-4 core is only 15mm and hence able to enclose a much smaller length of the current carrying conductor which is very essential in the power generation process. With the proposed core in Figure 4.2 of cross sectional areas $2cm^2$ and a height of 40mm, a lot more energy can be harvested. This core size will also fit in most boxes of electrical nodes.

An estimate of the amount of the magnetic flux densities around current carrying conductors in the built environment for power generation show an average of 0.25 T can be scavenged at a radial distance of 8mm from a conductor carrying 20A. An equivalent of 3VAC can be generated from this magnetic flux density. This means that for high current drawing nodes in the building electrical system, a much higher voltage can be induced with the proposed energy harvesting device. The material property of the proposed core has a saturation flux density of 1.56 T which is much higher to prevent the core from saturating.

The prototype board used in the experiments contained a typical electrical wiring system design. The loads used for testing were incandescent lamps and small size (wattage) heaters. To utilize power generated at all times, a rechargeable battery is included in the board setup to store energy when power is not being used by the energy management sensor. This board also displays an energy management sensor alongside LEDs connected to output power from the energy harvesting device. The next chapter looks at some of the technical challenges as well as potential future applications of this research.

Chapter 6

TECHNICAL CHALLENGES, BENEFITS AND OPPORTUNITIES

6.1 Introduction

Originally, electrical node boxes were designed to contain electrical cables and splice connectors. Node boxes come in various sizes as shown in NEC table 314.6(A) in Figure 4.1. With the exception of the node boxes, most electrical distribution cables are hidden in enclosed spaces within the built environment. Even though the proposed harvesting device is intended to supply power to energy management sensors in the built environment, other low powered devices can also be supplied with similar harvested power. This chapter examines some of the technical challenges associated with the design of the harvesting device, benefits that may be derived as a result of its use, as well as related future opportunities.

6.2 Challenges

Article 314.71 of the National Electrical Code (NEC) (52) states that pull and junction boxes and handhole enclosures shall provide adequate space and dimensions for conductor installation. Installing energy harvesting devices to power the sensor devices in the boxes of the nodes will reduce the allowable space required by the NEC.

Moreover, in order to power other low powered devices that are not necessarily located at the electrical nodes, access to current carrying conductors in the conduits will

be a limitation to scavenging magnetic flux density. Another limitation is mounting of harvesting device on current carrying conductors. That is, in order to scavenge for magnetic flux, the current carrying conductor must be mounted through the opening in the harvesting device (unlike other current transformers which can be clipped on to the conductor).

In applications where the energy demand of the sensors is less than the quantity generated, energy storage devices (chargeable batteries) have to be incorporated in the installation setup. If the energy scavenged is of an amount lower than the required power, then a back-up chargeable battery must be added to supplement the power requirement. This battery can be recharged with scavenged power during periods in which the energy management sensor is idle.

6.3 Benefits

This research has the potential to create a much more sustainable built environment and thus greatly benefit society. Because the energy source (magnetic field) is in close proximity to the energy management sensor, generated power can be directly applied to the sensor at its location. The ultimate goal of replacing batteries or supplying power to the hardwired energy management sensors with free perpetual energy can be achieved. This will save money in terms of the cost of batteries and all but eliminate the labor costs associated with battery replacement. Even with the limited amount of energy harvested, rechargeable batteries can be used in place of conventional batteries. Power generated by the energy harvesting device will continuously replenish used up energy from the batteries and thus prolong life of batteries.

Furthermore, electricity usage is ultimately maximized since there is additional capture when current is drawn through existing conductors. Thus, the magnetic field located around the current carrying conductors is put to efficient use.

Finally, one of the great advantages of using this device is that it can serve as a potential energy supply for common low powered devices such as motion sensors, phone chargers and some wall clocks in the built environment. Recent power outlet technologies (54), (55), (56) that feature a USB outlet for powering low powered devices up to 5VDC can be powered from this energy harvesting device when that node is drawing power. However, in this case it should be noted that the node will have to draw current in excess of 10A. Wall clocks typically mounted on hallway corridors walls can be powered with an energy harvesting device installed on current carrying conductors powering the lamps in the hallway. In order to constantly supply power to such a clock, a rechargeable battery backup can be incorporated and replenished from time to time by the harvesting device. Motion sensors in corridor hallways that are hardwired or battery operated can also be connected in a similar configuration.

6.4 Opportunities

The proposed energy harvesting device is potentially useful in a variety of other applications than the ones stated in this thesis. For example, if the device is permanently incorporated into the node boxes of the electrical system, power can be harvested for USB outlets. These outlets are currently being integrated into some newer power outlets (57). Therefore, the proposed energy harvesting device would therefore eliminate the need to hardwire these USB outlets to the existing electrical system.

This harvesting device may also be developed into an addressable sensor. That is, the harvesting device can be integrated with a transceiver circuit which will notify a parent transceiver whenever current is drawn at a node. The transceiver integrated with the harvesting device has a unique identification code that can also be used as an alerting system.

Finally, these harvesting devices may be mounted around high current drawing conductors. The harvested power could be channeled to a battery bank to power emergency lighting systems during power outages or other emergency situations.

The final chapter summarizes the whole research and presents the conclusion and recommendations for future work.

Chapter 7

CONCLUSION AND RECOMMENDATIONS

Four objectives were accomplished in this research. First, available magnetic flux was estimated around common current carrying conductors in the built environment. The Biot-Savart law's theoretical equation for estimating the magnetic flux of a given current flowing through a conductor is used to achieve this objective. Second, a comparison of the simulated induced voltages to the experimentally measured values proved accurate Faraday's law of induction. Using theoretical equations for current transformers derived from Faraday's law of induction, an estimate for the average voltages that can be induced from the magnetic flux on the current carrying conductors is given. Third, a harvesting device is designed that is small enough to fit into typical nodes of boxes in the electrical power distribution system in the built environment. It should be noted that due to manufacturing lead time, the prototype harvesting device used for experimenting and simulation is not of the proposed design shape. Thus, the geometry of the harvesting device used for experiments will not fit into a typical node box. Finally, a board was built for the purpose of experimenting with the prototype energy harvesting device. Experimental measurements reveal that an average of 2 V - 3V rms (equivalent short circuit current , 20 mA -40mA) can be induced from magnetic flux around current conductors in the built environment.

Based on a comparison of the experimental and simulated values, it is recommended that a prototype of the proposed design of the harvesting device be built and tested.

7.1 Future Research

Electromagnetic harvesting depends on a secondary ambient source of energy. Thus in order to harvest energy through the electromagnetic converting process, a changing magnetic flux density field must first be established. In the proposed researched, this field is generated when alternating current flows through electrical conductors in the built environment. In future research, investigations will be done to find out whether the harvesting operation affects the magnitude of current flowing through the primary conductor. A further investigation will be done to see if the type of primary load (resistive, inductive, capacitive) at a node in the electrical system has any relationship with magnetic flux density generation.

Another future research will be to investigate how much energy that can be scavenged from magnetic flux density in a given space volume. This investigation will also include harvesting electromagnetic energy around metallic conduits carrying electrical conductors in the built environment.

APPENDICES

Appendix A

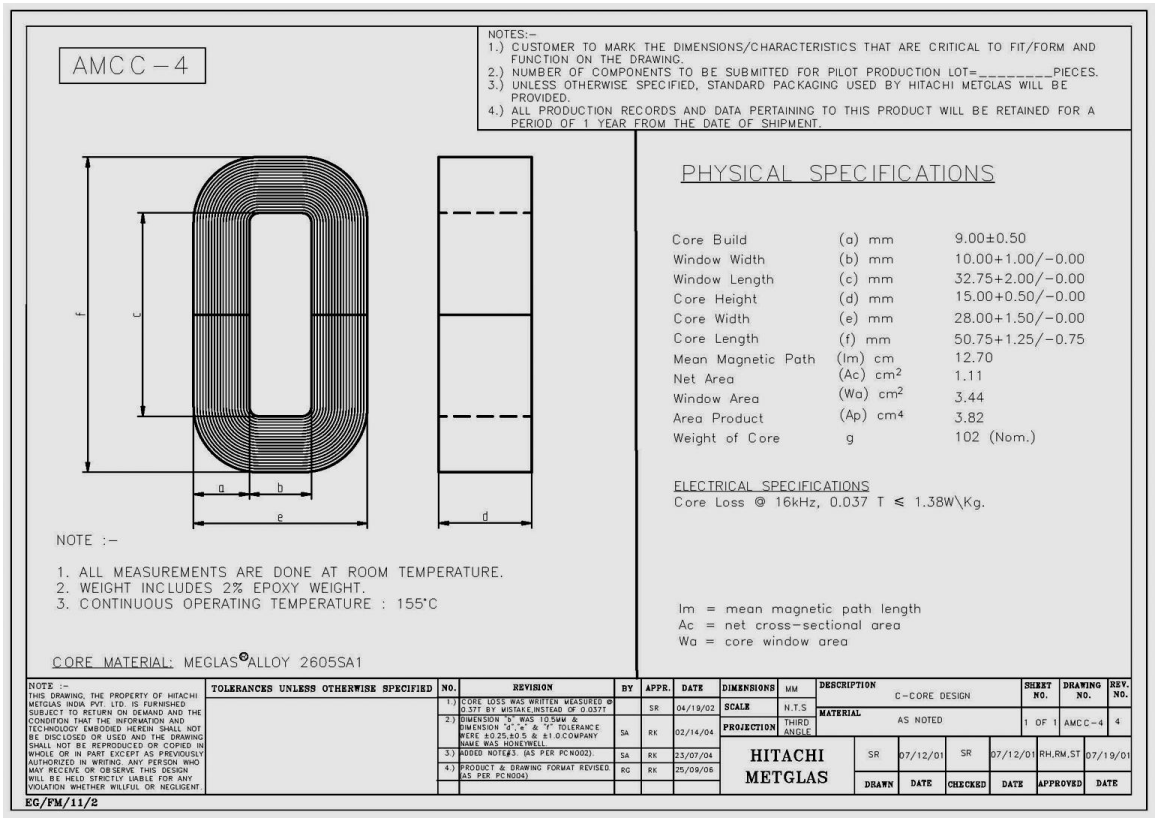
Appendix A.1: Wire Properties Table for Winding Conductors (50)

Wire Table

AWG	Bare Area		Resistance	Heavy Synthetics								
				Area		Diameter		Turns-Per		Turns-Per		Weight
	$cm^2(10^{-3})$	cir-mil	μΩ/cm 20°C	$cm^2(10^{-3})$	cir-mil	cm	Inch	cm	Inch	cm^2	$Inch^2$	gm/cm
1	2	3	4	5	6	7	8	9	10	11	12	13
10	52.6100	10384.00	32.7	55.9000	11046.00	0.2670	0.105	3.9	10	11	69	0.46800
11	41.6800	8226.00	41.4	44.5000	8798.00	0.2380	0.094	4.4	11	13	90	0.37500
12	33.0800	6529.00	52.1	35.6400	7022.00	0.2130	0.084	4.9	12	17	108	0.29770
13	26.2600	5184.00	65.6	28.3600	5610.00	0.1900	0.075	5.5	13	21	136	0.23670
14	20.8200	4109.00	82.8	22.9500	4556.00	0.1710	0.068	6.0	15	26	169	0.18790
15	16.5100	3260.00	104.3	18.3700	3624.00	0.1530	0.060	6.8	17	33	211	0.14920
16	13.0700	2581.00	131.8	14.7300	2905.00	0.1370	0.054	7.3	19	41	263	0.11840
17	10.3900	2052.00	165.8	11.6800	2323.00	0.1220	0.048	8.2	21	51	331	0.09430
18	8.2280	1624.00	209.5	9.3260	1857.00	0.1090	0.043	9.1	23	64	415	0.07474
19	6.5310	1289.00	263.9	7.5390	1490.00	0.0980	0.039	10.2	26	80	515	0.05940
20	5.1880	1024.00	332.3	6.0650	1197.00	0.0879	0.035	11.4	29	99	638	0.04726
21	4.1160	812.30	418.9	4.8370	954.80	0.0785	0.031	12.8	32	124	800	0.03757
22	3.2430	640.10	531.4	3.8570	761.70	0.0701	0.028	14.3	36	156	1003	0.02965
23	2.5880	510.80	666.0	3.1350	620.00	0.0632	0.025	15.8	40	191	1234	0.02372
24	2.0470	404.00	842.1	2.5140	497.30	0.0566	0.022	17.6	45	239	1539	0.01884
25	1.6230	320.40	1062.0	2.0020	396.00	0.0505	0.020	19.8	50	300	1933	0.01498
26	1.2800	252.80	1345.0	1.6030	316.80	0.0452	0.018	22.1	56	374	2414	0.01185
27	1.0210	201.60	1687.0	1.3130	259.20	0.0409	0.016	24.4	62	457	2947	0.00945
28	0.8046	158.80	2142.0	1.0515	207.30	0.0366	0.014	27.3	69	571	3680	0.00747
29	0.6470	127.70	2664.0	0.8548	169.00	0.0330	0.013	30.3	77	702	4527	0.00602
30	0.5067	100.00	3402.0	0.6785	134.50	0.0294	0.012	33.9	86	884	5703	0.00472
31	0.4013	79.21	4294.0	0.5596	110.20	0.0267	0.011	37.5	95	1072	6914	0.00372
32	0.3242	64.00	5315.0	0.4559	90.25	0.0241	0.010	41.5	105	1316	8488	0.00305
33	0.2554	50.41	6748.0	0.3662	72.25	0.0216	0.009	46.3	118	1638	10565	0.00241
34	0.2011	39.69	8572.0	0.2863	56.25	0.0191	0.008	52.5	133	2095	13512	0.00189
35	0.1589	31.36	10849.0	0.2268	44.89	0.0170	0.007	58.8	149	2645	17060	0.00150
36	0.1266	25.00	13608.0	0.1813	36.00	0.0152	0.006	62.5	167	3309	21343	0.00119
37	0.1026	20.25	16801.0	0.1538	30.25	0.0140	0.006	71.6	182	3901	25161	0.00098
38	0.0811	16.00	21266.0	0.1207	24.01	0.0124	0.005	80.4	204	4971	32062	0.00077
39	0.0621	12.25	27775.0	0.0932	18.49	0.0109	0.004	91.6	233	6437	41518	0.00059
40	0.0487	9.61	35400.0	0.0723	14.44	0.0096	0.004	103.6	263	8298	53522	0.00046
41	0.0397	7.84	43405.0	0.0584	11.56	0.0086	0.003	115.7	294	10273	66260	0.00038
42	0.0317	6.25	54429.0	0.0456	9.00	0.0076	0.003	131.2	333	13163	84901	0.00030
43	0.0245	4.84	70308.0	0.0368	7.29	0.0069	0.003	145.8	370	16291	105076	0.00023
44	0.0202	4.00	85072.0	0.0316	6.25	0.0064	0.003	157.4	400	18957	122272	0.00020

Appendix B

Appendix B.1: Specifications of AMCC-4 Core



AMCC – 4

NOTES:–
1.) CUSTOMER TO MARK THE DIMENSIONS/CHARACTERISTICS THAT ARE CRITICAL TO FIT/FORM AND FUNCTION ON THE DRAWING.
2.) NUMBER OF COMPONENTS TO BE SUBMITTED FOR PILOT PRODUCTION LOT=_____PIECES.
3.) UNLESS OTHERWISE SPECIFIED, STANDARD PACKAGING USED BY HITACHI METGLAS WILL BE PROVIDED.
4.) ALL PRODUCTION RECORDS AND DATA PERTAINING TO THIS PRODUCT WILL BE RETAINED FOR A PERIOD OF 1 YEAR FROM THE DATE OF SHIPMENT.

PHYSICAL SPECIFICATIONS

Core Build	(a) mm	9.00±0.50
Window Width	(b) mm	10.00+1.00/–0.00
Window Length	(c) mm	32.75+2.00/–0.00
Core Height	(d) mm	15.00+0.50/–0.00
Core Width	(e) mm	28.00+1.50/–0.00
Core Length	(f) mm	50.75+1.25/–0.75
Mean Magnetic Path	(lm) cm	12.70
Net Area	(Ac) cm^2	1.11
Window Area	(Wa) cm^2	3.44
Area Product	(Ap) cm^4	3.82
Weight of Core	g	102 (Nom.)

ELECTRICAL SPECIFICATIONS
Core Loss @ 16kHz, 0.037 T ≤ 1.38W\Kg.

NOTE :–
1. ALL MEASUREMENTS ARE DONE AT ROOM TEMPERATURE.
2. WEIGHT INCLUDES 2% EPOXY WEIGHT.
3. CONTINUOUS OPERATING TEMPERATURE : 155°C

lm = mean magnetic path length
Ac = net cross–sectional area
Wa = core window area

CORE MATERIAL: MEGLAS® ALLOY 2605SA1

NOTE :–
THIS DRAWING, THE PROPERTY OF HITACHI METGLAS INDIA PVT. LTD. IS FURNISHED SUBJECT TO RETURN ON DEMAND AND THE CONDITION THAT THE INFORMATION AND TECHNOLOGY EMBODIED HEREIN SHALL NOT BE DISCLOSED OR USED AND THE DRAWING SHALL NOT BE REPRODUCED OR COPIED IN WHOLE OR IN PART EXCEPT AS PREVIOUSLY AUTHORIZED IN WRITING. ANY PERSON WHO MAY RECEIVE OR OBSERVE THIS DESIGN WILL BE HELD STRICTLY LIABLE FOR ANY VIOLATION WHETHER WILLFUL OR NEGLIGENT.

TOLERANCES UNLESS OTHERWISE SPECIFIED

NO.	REVISION	BY	APPR.	DATE
1.)	CORE LOSS WAS WRITTEN MEASURED @ 0.37T BY MISTAKE,INSTEAD OF 0.037T		SR	04/19/02
2.)	DIMENSION "b" WAS 10.5MM & DIMENSION "d","e" & "f" TOLERANCE WERE ±0.25,±0.5 & ±1.0 COMPANY NAME WAS HONEYWELL	SA	RK	02/14/04
3.)	ADDED NOTE#3. (AS PER PCN002).	SA	RK	23/07/04
4.)	PRODUCT & DRAWING FORMAT REVISED (AS PER PCN004)	RG	RK	25/09/08

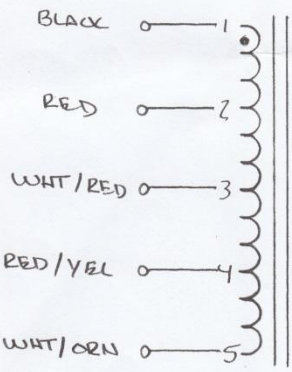
DIMENSIONS	MM	DESCRIPTION	C–CORE DESIGN	SHEET NO.
SCALE	N.T.S			DRAWING NO.
PROJECTION	THIRD ANGLE	MATERIAL	AS NOTED	REV. NO.

HITACHI METGLAS	SR	07/12/01	SR	07/12/01	RH,RM,ST	07/19/01	1 OF 1 AMCC–4 4
	DRAWN	DATE	CHECKED	DATE	APPROVED	DATE	

EG/FM/11/2

Appendix B.2: Winding Schematic Drawings of AMCC-4 Core

FROST MAGNETICS INC



PART NO. FM-8361-A
DATE 2-13-12 ENG. MZ
CUST. EVANS
CUST. NO. SAME
CU. LOSS
FE LOSS
TOTAL LOSSES
VA / LB I_{EX}
SA
ΔT

CORE AMCC-4			WND	LEG	A_C (NET)	LC	VOL	FLUX DENS.	W/LB	WT (NET
WDG. FORM 2-L-KAPTON	PR.									
TERMINAL										
LEAD NO.	1- 2-	3-4-5								
VOLTAGE (O.C.)										
CURRENT										
WIRE	#25 SN									
O.D.										
C.M.										
CM/A										
WL										
T/L										
LAYERS										
TURNS	0→200→	300→400→500								
TAPS										
LAYER INS.										
WRAPPER	1-L DACRON									
FT/TURN										
TOTAL LENGTH										
OHMS/M										
DCR										
IR/% REG										
LB/M										
WEIGHT										
I^2R										
BUILD										

BUILD UP

Appendix C

Appendix C.1: MATLAB Programming Codes

```
%*********************************************************************
%%%%%%%%%% This code was written by Evans Sordiashie %%%%%%%%%%%%%%%%
%*********************************************************************
%% MATLAB Code developed for Master of Science Thesis, May 2012, UNL.%%

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
Title: Electromagnetic Harvesting to power Energy Management Sensors
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%

%This MATLAB code is used to determine the following:

% 1. Plot Magnetic Field around a Current carrying conductor

%*********************************************************************
clear all
close all
clc
%====================================================================
% current carrying conductor conductor parameters
%====================================================================
L = 100;                    % length of conductor
mu0 = 4*pi*1e-7;     % magnetic permeability constant
N=100;                    % number of elements in current carrying
conductor
I=10;                     % current flowing through the conductor (A)
%====================================================================
xC = zeros(1,N);       % xyz coordinates points on current conductor
yC = zeros(1,N);
zC = zeros(1,N);

%====================================================================
%Discretizing conductor
%====================================================================
 dl=L/N; l(1)=dl/2;

for i=2:N,
    l(i) = (i-1)*dl + l(1);
end

zC = l;

% length of each elemental conductor element
Lx = 0; Ly = 0; Lz= dl;

% Detector space(rectangular coordinates~xP, yP, zP) where B is
calculated
NP = 200;                     % Detector points NP x N
xPmax = 100;                  % Dimensions of detector space
```

```
yPmin = 0; yPmax = N;

xP = linspace(-xPmax,xPmax,NP);
yP = linspace(-yPmax,yPmax,NP);
[xxP yyP] = meshgrid(xP,yP);
Bx = zeros(NP,NP);By = Bx; Bz = Bx;

%=========================================================================
% Calculation of magnetic field B
%=========================================================================
for c = 1 : N
   rx =  xxP;
   ry = yyP;
   rz = zC(c);
   r  = sqrt(rx.^2 + ry.^2 + rz.^2);
   r2 = r.^2;
   Bx = Bx + (mu0*I./(4*pi))*Lz.*ry./r2;
   By = By + (mu0*I./(4*pi))*Lz.*rx./r2;
   Bz = Bz + 0;
end
B = sqrt(Bx.^2 + By.^2 + Bz.^2);
B = B./max(max(B));            % normalize B to 1

%=========================================================================
% Simulation of Magnetic field for given current I
%=========================================================================
%2-D Plot
%========
figure(1)
pcolor(xxP,yyP,B);
colormap(jet)
shading interp;
axis equal;
axis([-xPmax xPmax 0 yPmax]);
xlabel('Center of Conductor','FontWeight','bold','FontSize',12);
ylabel('Radial Distances from center of
conductor','FontWeight','bold'....
    ,'FontSize',12);
set(gca,'XTick',[0:50:100]);
set(gca,'YTick',[0:50:100]);
rectangle('Position',[-1 0 2 0.2],'FaceColor','k');
title('MagneticFlux Density Around a 10A Current Carrying
Conductor',...
    'FontWeight','bold','FontSize',14)
h = colorbar('horiz');
set(h, 'XTick')
xlabel(h, 'Tesla')
%=========================================================================
%3-D View
%========
figure(2)

surf(xxP,yyP,B,'FaceColor','interp',...
    'EdgeColor','none',...
    'FaceLighting','phong')
daspect([1 1 1])
```

```
axis tight
xlabel('Distances from current carrying
conductor','FontWeight','bold'...
    ,'FontSize',12,'Rotation',-44);
ylabel('Distances from current carrying conductor',
'FontWeight','bold'...
    ,'FontSize',12,'Rotation',19.5);
view(-122,36)
camlight left
colormap(jet)
% colorbar
title('Magnetic Flux Density Around a 10A Current Carrying
Conductor',...
    'FontWeight','bold','FontSize',14)
h = colorbar;
ylabel = get(h,'YTickLabel');
T = repmat('T',length(ylabel),1);
ylabel = [ylabel T];
set(h,'YTickLabel',ylabel);

%=========================================================================
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% End %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
%=========================================================================

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
%Title: Electromagnetic Harvesting to power Energy Management Sensors%%
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%

%This MATLAB code is used to determine the following:

% 1. Magnetic field strength around a conductor carrying 1-20A
% 2. Plot of magnetic field strength against currents
% 3. Plot of Induced voltage vrs primary conductor current (1-20A)
% 4. Plot of DC voltage vrs primary conductor current (1-20A)

%*************************************************************************
% Electromagnetic Harvesting to power Energy Management Sensors

clear all
close all
clc

% parameters for current carrying conductor
L = 15; % length (mm) copper #12 AWG conductor
C = 2.05/2; % conductor average radius (mm), NEC Table 8

%=========================================================================
% Core properties
%=================
lm = 12.2e-2; %(m), Magnetic path length (from core data), AMCC-4 Core
lg = 120e-6; %(m), Air gap between cut cores(range 25-150 micro meter)
mu_m = 3000; % (H/m), permeability of core material (Amorphous, 2605)
mur = mu_m/(1 + mu_m *(lg/(lm)))%; relative permeability of core
mu0 = 4e-7 * pi; % Magnetic permeability of free space or vacuum
mu = mu0 * mur;% (H/m), Effective permeability of core
```

```
%=========================================================================
% Calculating Magnetic Strength around current carrying conductor
%=========================================================================
Ip = 1:1:20; % Primary current range
I = Ip;
r = (C:.025:30);% radial distances from current carrying
conductor,0.025 increments

B = zeros(1160,1); % Magnetic field Strength

for  p =1:20;
    for i = 1:1160;
        B(i,p) = (mu * I(p) * L * 1e-3)/(2*pi*r(i)*1e-3 *...
            sqrt(4*(r(i)*1e-3)^2 + (L*1e-3)^2));
    end
end


r1 =r(41:1160)'; % Selected radial distances
Bac = B(41:1160,1:20); % Corresponding magnetic field strength
H = [r1 B(41:1160,1:20)]; % Matrix of radial distances and magnetic
flux
%=========================================================================
% Plotting generated magnetic field against radial distance. Field
% generated for 1-20A
%=========================================================================
clf
figure (1)
plot(r1, Bac(:,1:20));
set(gca,'XTick',2:2:30)
set(gca,'YTick',0:0.1:1.5)
grid on
xlabel({'Distance from current carrying primary conductor (mm)'},...
    'FontWeight','bold','FontSize',12);
ylabel({'Magnetic Flux Density
(T)'},'FontWeight','bold','FontSize',12);
title('Magnetic Flux Density Around #12 AWG Current Carrying Conductor
(1-20A)',...
    'FontWeight','bold',...
    'FontSize',12);
%Create arrow
annotation1 = annotation(...
  'arrow',...
  [0.1795 0.312],[0.1441 0.3594],...
  'LineWidth',2);

% Create textbox
annotation2 = annotation(...
  'textbox',...
  'Position',[0.306 0.3363 0.2096 0.05516],...
  'LineStyle','none',...
  'FitHeightToText','off',...
  'FontSize',12,...
```

```
  'FontWeight','bold',...
  'String',{'Order of increasing current(1-20A)'});


%=========================================================================
% Calculating induced Voltage
%=========================================================================
% Core Parameters
% %===============
Ac = 1.11;% Cross sectional area of core in cm.sq, AMCC-4 Core.

%Operating Parameters
%====================
Np = 1; % Number of turns of primary conductor ~1
f = 60; % output frequency in Hz
Kf = 4.44; % Waveform factor, sinewave.
%=========================================================================
% Magnetic field strength at (,6,7,8,9,10mm)from current carrying...
% conductor is used for calculation ~ approx. distance from magnetic
path length(lm or MPL)

D = [Bac(201,:)' Bac(241,:)' Bac(281,:)' Bac(321,:)' Bac(361,:)'];
%=========================================================================
% Calculating induced voltages for various number of magnetic wire
windings
% around core. Ns = 50, 100, 150, 200, 250, 300, 350, 400, 450, 500.
%=========================================================================
Ns = (50:50:500);
Vs = zeros(20,5);
for w = 1:5,
    for g = 1:20,
        for m = 1:10,
            Vs (g,w,m) = Ac * Kf * D(g,w) * f * Ns(m)/(1e4);
        end
    end
end
Vs;

%=========================================================================
%Plotting Induced Voltages against Primary Current (A) for 200, 300,
400, 500turns.
%=========================================================================

% figure (2)
jj = figure('Color',[1 1 1]);
subplot(2,2,1)
plot1=plot (Ip, Vs(:,:,4));
set(plot1(1),'LineWidth',1.0,'Marker','o');
set(plot1(2),'LineWidth',1.0,'Marker','square')
set(plot1(3),'LineWidth',2.0,'LineStyle','--');
set(plot1(4),'LineWidth',1.0,'Marker','diamond');
set(plot1(5),'LineWidth',1.0,'Marker','^');
set(gca,'XTick',1:1:20)
grid on
xlabel({'Primary Current (A)'},'FontWeight','bold','FontSize',12);
```

```
ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Current (200 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
legend('6mm from MPL','7mm from MPL','8mm from MPL','9mm from
MPL','10mm from MPL')

set(legend,...
    'Position',[0.146979166666665 0.750053049452929 0.113125
0.144132653061225],...
    'FontSize',10);

subplot(2,2,2)
plot1=plot (Ip, Vs(:,:,6));
set(plot1(1),'LineWidth',1.0,'Marker','o');
set(plot1(2),'LineWidth',1.0,'Marker','square')
set(plot1(3),'LineWidth',2.0,'LineStyle','--');
set(plot1(4),'LineWidth',1.0,'Marker','diamond');
set(plot1(5),'LineWidth',1.0,'Marker','^');
set(gca,'XTick',1:1:20)
grid on
xlabel({'Primary Current (A)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Current (300 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
legend('6mm from MPL','7mm from MPL','8mm from MPL','9mm from
MPL','10mm from MPL')

set(legend,'Position',[0.602916666666662 0.758343865779456 0.09625
0.125]);

subplot(2,2,3)
plot1=plot (Ip, Vs(:,:,8));
set(plot1(1),'LineWidth',1.0,'Marker','o');
set(plot1(2),'LineWidth',1.0,'Marker','square')
set(plot1(3),'LineWidth',2.0,'LineStyle','--');
set(plot1(4),'LineWidth',1.0,'Marker','diamond');
set(plot1(5),'LineWidth',1.0,'Marker','^');
set(gca,'XTick',1:1:20)
grid on
xlabel({'Primary Current (A)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Current (400 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
legend('6mm from MPL','7mm from MPL','8mm from MPL','9mm from
MPL','10mm from MPL')

set(legend,'Position',[0.15916666666666 0.282578559657006 0.09625
0.125]);

subplot(2,2,4)
```

```
plot1=plot (Ip, Vs(:,:,10));
set(plot1(1),'LineWidth',1.0,'Marker','o');
set(plot1(2),'LineWidth',1.0,'Marker','square')
set(plot1(3),'LineWidth',2.0,'LineStyle','--');
set(plot1(4),'LineWidth',1.0,'Marker','diamond');
set(plot1(5),'LineWidth',1.0,'Marker','^');
set(gca,'XTick',1:1:20)
grid on
xlabel({'Primary Current (A)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Current (500 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
legend('6mm from MPL','7mm from MPL','8mm from MPL','9mm from
MPL','10mm from MPL')

set(legend,'Position',[0.609166666666654 0.294058151493739 0.09625
0.125]);

ha = axes('Position',[0 0 1 1],'Xlim',[0 1],'Ylim',[0
1],'Box','off','Visible','off','Units','normalized', 'clipping' ,
'off');

text(0.5, 1,'\bf Induced Voltage with AMCC - 4
Core','HorizontalAlignment'...
,'center','VerticalAlignment', 'top','FontWeight','bold',...
    'FontSize',14);

%=========================================================================
%Calculating output DC Voltage
%=========================================================================
Vd = 0.5; % Diode voltage drop

Vo = zeros(20,5);
for k = 1:5,
    for t = 1:20,
        for j = 1:10,
            Vo(t,k,j) = Vs(t,k,j) - Vd;
        end
    end
end
Vo;

%=========================================================================
%Plotting DC Voltages against Primary Current (A) for 200, 300, 400,
%500 turns
%=========================================================================

figure (3)
subplot(2,2,1)
plot1=plot (Ip, Vo(:,:,4));
set(plot1(1),'LineWidth',1.0,'Marker','o');
set(plot1(2),'LineWidth',1.0,'Marker','square')
set(plot1(3),'LineWidth',2.0,'LineStyle','--');
```

```
set(plot1(4),'LineWidth',1.0,'Marker','diamond');
set(plot1(5),'LineWidth',1.0,'Marker','^');
set(gca,'XTick',1:1:20)
grid on
xlabel({'Primary Current (A)'},'FontWeight','bold','FontSize',12);

ylabel({'DC Voltage (V)'},'FontWeight','bold','FontSize',12);

title({'DC Voltage Vrs Primary Current (200 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
legend('6mm from MPL','7mm from MPL','8mm from MPL','9mm from
MPL','10mm from MPL')

set(legend,...
    'Position',[0.146979166666665 0.750053049452929 0.113125
0.144132653061225],...
    'FontSize',10);

subplot(2,2,2)
plot1=plot (Ip, Vo(:,:,6));
set(plot1(1),'LineWidth',1.0,'Marker','o');
set(plot1(2),'LineWidth',1.0,'Marker','square')
set(plot1(3),'LineWidth',2.0,'LineStyle','--');
set(plot1(4),'LineWidth',1.0,'Marker','diamond');
set(plot1(5),'LineWidth',1.0,'Marker','^');
set(gca,'XTick',1:1:20)
grid on
xlabel({'Primary Current (A)'},'FontWeight','bold','FontSize',12);

ylabel({'DC Voltage (V)'},'FontWeight','bold','FontSize',12);

title({'DC Voltage Vrs Primary Current (300 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
legend('6mm from MPL','7mm from MPL','8mm from MPL','9mm from
MPL','10mm from MPL')

set(legend,'Position',[0.602916666666662 0.758343865779456 0.09625
0.125]);

subplot(2,2,3)
plot1=plot (Ip, Vo(:,:,8));
set(plot1(1),'LineWidth',1.0,'Marker','o');
set(plot1(2),'LineWidth',1.0,'Marker','square')
set(plot1(3),'LineWidth',2.0,'LineStyle','--');
set(plot1(4),'LineWidth',1.0,'Marker','diamond');
set(plot1(5),'LineWidth',1.0,'Marker','^');
set(gca,'XTick',1:1:20)
grid on
xlabel({'Primary Current (A)'},'FontWeight','bold','FontSize',12);

ylabel({'DC Voltage (V)'},'FontWeight','bold','FontSize',12);

title({'DC Voltage Vrs Primary Current (400 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
```

```
legend('6mm from MPL','7mm from MPL','8mm from MPL','9mm from
MPL','10mm from MPL')

set(legend,'Position',[0.15916666666666 0.282578559657006 0.09625
0.125]);

subplot(2,2,4)
plot1=plot (Ip, Vo(:,:,10));
set(plot1(1),'LineWidth',1.0,'Marker','o');
set(plot1(2),'LineWidth',1.0,'Marker','square')
set(plot1(3),'LineWidth',2.0,'LineStyle','--');
set(plot1(4),'LineWidth',1.0,'Marker','diamond');
set(plot1(5),'LineWidth',1.0,'Marker','^');
set(gca,'XTick',1:1:20)
grid on
xlabel({'Primary Current (A)'},'FontWeight','bold','FontSize',12);

ylabel({'DC Voltage (V)'},'FontWeight','bold','FontSize',12);

title({'DC Voltage Vrs Primary Current (500 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
legend('6mm from MPL','7mm from MPL','8mm from MPL','9mm from
MPL','10mm from MPL')

set(legend,'Position',[0.609166666666654 0.294058151493739 0.09625
0.125]);

ha = axes('Position',[0 0 1 1],'Xlim',[0 1],'Ylim',[0
1],'Box','off','Visible','off','Units','normalized', 'clipping' ,
'off');

text(0.5, 1,'\bf DC Voltage with AMCC - 4
Core','HorizontalAlignment'...
,'center','VerticalAlignment', 'top','FontWeight','bold',...
    'FontSize',14);

%=========================================================================
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
%=========================================================================

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
%%Title: Electromagnetic Harvesting to power Energy Management Sensors%
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%

%This MATLAB code is used to determine the following:

% 1. Plot of Induced Voltage (Vrms) against Primary Load 100-2400(W)
% 2. Plot of Induced voltage vrs Primary Load vrs radial distance
%*************************************************************************
 clear all
close all
clc

% parameters for current carrying conductor
```

```
L = 15; % length (mm) copper #12 AWG conductor
C = 2.05/2; % conductor average radius (mm), NEC Table 8

%=========================================================================
% Core properties
%=================
lm = 12.2e-2; %(m), Magnetic path length (from core data), AMCC-4 Core
lg = 120e-6; %(m), Air gap between cut cores(range 25-150 micro meter)
mu_m = 3000; % (H/m), permeability of core material (Amorphous, 2605)
mur = mu_m/(1 + mu_m *(lg/(lm))); %relative permeability of core
mu0 = 4e-7 * pi; % Magnetic permeability of free space or vacuum
mu = mu0 * mur;% (H/m), Effective permeability of core


%=========================================================================
% Calculating Magnetic Strength around current carrying conductor
%=========================================================================
% Primary Load (Watts)
Rp = 100:100:2400;

%Primary Voltage (V rms)
Vp = 120;

Ip = Rp/Vp;

r = (C:.025:30);% radial distances from current carrying
conductor,0.025 increments

B = zeros(1160,1);
for  p =1:24;
    for i = 1:1160;
        B(i,p) = (mu * Ip(p) * L * 1e-3)/(2*pi*r(i)*1e-3 *...
            sqrt(4*(r(i)*1e-3)^2 + (L*1e-3)^2));
    end
end

%=========================================================================
% Selecting magnetic field strength values at radial distances from...
% current carrying conductor (1, 2, 3, 4, 5, 6, 7, 8, 9, 10mm ).
Current...
% range for #12 AWG conductor is 1 to 20A.
%=========================================================================

r1 =r(41:1160)'; % radial distances
Bac = B(41:1160,1:24); % Corresponding magnetic field strength
% H = [r1 B(41:1160,1:24)]; % Matrix of radial distances and magnetic
flux

%=========================================================================
% Calcuting Induced voltage//Secondary turns//Output Voltage
%=========================================================================

% Core Parameters
%====================
Ac = 1.11;% Cross sectional area of core in cm.sq, AMCC-4 Core.
```

```
%Operating Parameters
%====================
Np = 1; % Number of turns of primary conductor ~1
f = 60; % output frequency in Hz
Kf = 4.44; % Waveform factor, sinewave.

% Magnetic field strength at (,6,7,8,9,10mm)from current carrying...
% conductor is used for calculation ~ approx distance from magnetic
path...
% length(lm or MPL)
%=========================================================================
D = [Bac(201,:)' Bac(241,:)' Bac(281,:)' Bac(321,:)' Bac(361,:)'];

%=========================================================================
% Calculating induced voltages for various number of magnetic wire
windings
% around core. Ns = 50, 100, 150, 200, 250, 300, 350, 400, 450, 500.
%=========================================================================
Ns = (50:50:500);
Vs = zeros(24,5);
for w = 1:5,
    for g = 1:24,
        for m = 1:10,
            Vs (g,w,m) = Ac * Kf * D(g,w) * f * Ns(m)/(1e4);
        end
    end
end
Vs%;

%=========================================================================
%Plotting Induced Voltages against Primary Load (W) for 200, 300, 400,
%500... turns
%=========================================================================
clf
figure (1)
% jp = figure('Color',[1 1 1]);
subplot(2,2,1)
plot1=plot (Rp, Vs(:,:,4));
set(plot1(1),'LineWidth',1.0,'Marker','o');
set(plot1(2),'LineWidth',1.0,'Marker','square')
set(plot1(3),'LineWidth',2.0,'LineStyle','--');
set(plot1(4),'LineWidth',1.0,'Marker','diamond');
set(plot1(5),'LineWidth',1.0,'Marker','^');
set(gca,'XTick',200:200:2400)

grid on
xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (200 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
legend('6mm from MPL','7mm from MPL','8mm from MPL','9mm from MPL',...
    '10mm from MPL')
```

```
set(legend,...
    'Position',[0.146979166666665 0.750053049452929 0.113125...
    0.144132653061225],'FontSize',10);

subplot(2,2,2)
plot1=plot (Rp, Vs(:,:,6));
set(plot1(1),'LineWidth',1.0,'Marker','o');
set(plot1(2),'LineWidth',1.0,'Marker','square')
set(plot1(3),'LineWidth',2.0,'LineStyle','--');
set(plot1(4),'LineWidth',1.0,'Marker','diamond');
set(plot1(5),'LineWidth',1.0,'Marker','^');
set(gca,'XTick',200:200:2400)

grid on
xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (300 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
legend('6mm from MPL','7mm from MPL','8mm from MPL','9mm from
MPL','10mm from MPL')

set(legend,'Position',[0.602916666666662 0.758343865779456 0.09625
0.125]);

subplot(2,2,3)
plot1=plot (Rp, Vs(:,:,8));
set(plot1(1),'LineWidth',1.0,'Marker','o');
set(plot1(2),'LineWidth',1.0,'Marker','square')
set(plot1(3),'LineWidth',2.0,'LineStyle','--');
set(plot1(4),'LineWidth',1.0,'Marker','diamond');
set(plot1(5),'LineWidth',1.0,'Marker','^');
set(gca,'XTick',200:200:2400)

grid on
xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (400 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
legend('6mm from MPL','7mm from MPL','8mm from MPL','9mm from
MPL','10mm from MPL')

set(legend,'Position',[0.15916666666666 0.282578559657006 0.09625
0.125]);

subplot(2,2,4)
plot1=plot (Rp, Vs(:,:,10));
set(plot1(1),'LineWidth',1.0,'Marker','o');
set(plot1(2),'LineWidth',1.0,'Marker','square')
set(plot1(3),'LineWidth',2.0,'LineStyle','--');
```

```
set(plot1(4),'LineWidth',1.0,'Marker','diamond');
set(plot1(5),'LineWidth',1.0,'Marker','^');
set(gca,'XTick',200:200:2400)

grid on
xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (500 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
legend('6mm from MPL','7mm from MPL','8mm from MPL','9mm from
MPL','10mm from MPL')

set(legend,'Position',[0.609166666666654 0.294058151493739 0.09625
0.125]);

ha = axes('Position',[0 0 1 1],'Xlim',[0 1],'Ylim',[0
1],'Box','off','Visible','off','Units','normalized', 'clipping' ,
'off');

text(0.5, 1,'\bf Induced Voltage for AMCC - 4
Core','HorizontalAlignment'...
,'center','VerticalAlignment', 'top','FontWeight','bold',...
    'FontSize',14);

figure (2)
jpm = figure('Color',[1 1 1]);
subplot(2,2,1)
surf(Vs(:,:,4),'MarkerEdgeColor','none','LineWidth',0.1,...
    'LineStyle','none',...
    'FaceColor','interp',...
    'EdgeColor','none');
colorbar([0.0723362878896467 0.596376914225015 0.0076637121103533...
    0.340686022837922]);
set(gca,'XTick',1:1:5)
set(gca,'YTick',0:5:25)
set(gca,'ZTick',0:0.5:5)

xlabel({'Dists., D1-D5(6mm,7mm,8mm,9mm,and
10mm)'},'FontWeight','bold',...
    'FontSize',8,'Rotation',9);

ylabel({'Primary Load
(W)'},'FontWeight','bold','FontSize',8,'Rotation',-16);

zlabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',8);

title({'Induced Voltage Vrs Primary Load (200 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',10);


subplot(2,2,2)
surf(Vs(:,:,6),'MarkerEdgeColor','none','LineWidth',0.1,...
```

```
    'LineStyle','none',...
    'FaceColor','interp',...
    'EdgeColor','none');
colorbar([0.937336287889647 0.573649641497743 0.00641371211035334...
    0.345930778082677]);
set(gca,'XTick',1:1:5)
set(gca,'YTick',0:5:25)
set(gca,'ZTick',0:0.5:5)

xlabel({'Dists., D1-D5(6mm,7mm,8mm,9mm,and
10mm)'},'FontWeight','bold',...
    'FontSize',12,'Rotation',9);

ylabel({'Primary Load
(W)'},'FontWeight','bold','FontSize',12,'Rotation',-16);

zlabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (300 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',14);

subplot(2,2,3)
surf(Vs(:,:,8),'MarkerEdgeColor','none','LineWidth',0.1,...
    'LineStyle','none',...
    'FaceColor','interp',...
    'EdgeColor','none');
colorbar([0.0717112878896468 0.106643356643357 0.00828871211035319...
    0.339160839160839]);
set(gca,'XTick',1:1:5)
set(gca,'YTick',0:5:25)
set(gca,'ZTick',0:0.5:5)

xlabel({'Dists., D1-D5(6mm,7mm,8mm,9mm,and
10mm)'},'FontWeight','bold',...
    'FontSize',8,'Rotation',9);

ylabel({'Primary Load
(W)'},'FontWeight','bold','FontSize',8,'Rotation',-16);

zlabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',8);

title({'Induced Voltage Vrs Primary Load (400 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',10);

subplot(2,2,4)
surf(Vs(:,:,10),'MarkerEdgeColor','none','LineWidth',0.1,...
    'LineStyle','none',...
    'FaceColor','interp',...
    'EdgeColor','none');
colorbar([0.937961287889648 0.134838452686554 0.00703871211035268...
    0.340686022837922]);
set(gca,'XTick',1:1:5)
set(gca,'YTick',0:5:25)
set(gca,'ZTick',0:0.5:5)
```

```
xlabel({'Dists., D1-D5(6mm,7mm,8mm,9mm,and
10mm)'},'FontWeight','bold',...
    'FontSize',12,'Rotation',9);

ylabel({'Primary Load
(W)'},'FontWeight','bold','FontSize',12,'Rotation',-16);

zlabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (500 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',14);

ha = axes('Position',[0 0 1 1],'Xlim',[0 1],'Ylim',[0
1],'Box','off','Visible','off','Units','normalized', 'clipping' ,
'off');

text(0.5, 1,'\bf Induced Voltage for AMCC - 4
Core','HorizontalAlignment'...
,'center','VerticalAlignment', 'top','FontWeight','bold',...
    'FontSize',14);

%=========================================================================
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%   End  %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
%=========================================================================



%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
%%Title: Electromagnetic Harvesting to power Energy Management Sensors%
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%

%This MATLAB code is used to determine the following:

% 1. Plot experimental measurements of induced Open Circuit Voltages
% 2. Plot experimental measurements of short circuit currents

%*************************************************************************
% Electromagnetic Harvesting to power Energy Management Sensors
clear all
close all
clc

% parameters for current carrying conductor
L = 15; % length (mm) copper #12 AWG conductor
C = 2.05/2; % conductor average radius (mm), NEC Table 8

%=========================================================================

Pl = [75; 150; 650; 725; 800; 1500; 1650; 2150; 2300];% Primary Load

% Induced voltage #25 AWG wire - 200 turns, AMCC-4 Core (Open circuit
% voltage, Vso, V rms) and short circuit current Iso (mA)

Vso_a = [0.03; 0.05; 0.34; 0.35; 0.40; 0.82; 0.88; 1.17; 1.23];
```

```
Iso_a = [3.02; 4.58; 24.88; 26.41; 29.64; 57.7; 62.2; 82.3; 86.9];

% Induced voltage #25 AWG wire - 300 turns, AMCC-4 Core (Open circuit
% voltage, Vso, V rms) and short circuit current Iso (mA)

Vso_b = [0.05; 0.08; 0.52; 0.56; 0.62; 1.27; 1.37; 1.82; 1.92];

Iso_b = [1.98; 3.02; 16.40; 17.36; 19.45; 37.80; 40.80; 53.90; 56.9];

% Induced voltage #25 AWG wire - 400 turns, AMCC-4 Core (Open circuit
% voltage, Vso, V rms) and short circuit current Iso (mA)

Vso_c = [0.07; 0.1; 0.7; 0.75; 0.84; 1.7; 1.83; 2.43; 2.55];

Iso_c = [1.49; 2.28; 12.42; 13.10; 14.67; 28.5; 30.8; 40.6; 42.9];

% Induced voltage #25 AWG wire - 500 turns, AMCC-4 Core (Open circuit
% voltage, Vso, V rms) and short circuit current Iso (mA)

Vso_d = [0.08;0.13; 0.89; 0.93; 1.05; 2.12; 2.28; 3.04; 3.19];

Iso_d = [1.20; 1.82; 10.09; 10.53; 11.80; 23.0; 24.7; 33.01; 34.5];

%Plots of Lab results
figure(1)
jpp = figure('Color',[1 1 1]);
subplot(2,2,1)
plot(Pl,Vso_a,'LineWidth',2)
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (200 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on

subplot(2,2,2)
plot(Pl,Iso_a,'LineWidth',2)
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Short Circuit Current
(mA)'},'FontWeight','bold','FontSize',12);

title({'Short Circuit Current Vrs Primary Load (200 turns, #25
AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on
subplot(2,2,3)
```

```
plot(Pl,Vso_b,'LineWidth',2)
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (300 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on
subplot(2,2,4)
plot(Pl,Iso_b,'LineWidth',2)
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Short Circuit Current
(mA)'},'FontWeight','bold','FontSize',12);

title({'Short Circuit Current Vrs Primary Load (300 turns, #25
AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on

ha = axes('Position',[0 0 1 1],'Xlim',[0 1],'Ylim',[0
1],'Box','off','Visible','off','Units','normalized', 'clipping' ,
'off');

text(0.5, 1,'\bf Experimental Induced Open Cct Voltage and Equivalent
Short cct Current with AMCC - 4 Core',...
    'HorizontalAlignment','center','VerticalAlignment', 'top',...
    'FontWeight','bold','FontSize',14);


figure(2)
jp = figure('Color',[1 1 1]);
subplot(2,2,1)
plot(Pl,Vso_c,'LineWidth',2)
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (400 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on
subplot(2,2,2)
plot(Pl,Iso_c,'LineWidth',2)
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);
```

```
ylabel({'Short Circuit Current
(mA)'},'FontWeight','bold','FontSize',12);

title({'Short Circuit Current Vrs Primary Load (400 turns, #25
AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on
subplot(2,2,3)
plot(Pl,Vso_d,'LineWidth',2)
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (500 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on
subplot(2,2,4)
plot(Pl,Iso_d,'LineWidth',2)
set(gca,'XTick',200:200:2400)
xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Short Circuit Current
(mA)'},'FontWeight','bold','FontSize',12);

title({'Short Circuit Current Vrs Primary Load (500 turns, #25
AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on


ha = axes('Position',[0 0 1 1],'Xlim',[0 1],'Ylim',[0
1],'Box','off','Visible','off','Units','normalized', 'clipping' ,
'off');

text(0.5, 1,'\bf Experimental Induced Open Cct Voltage and Equivalent
Short cct Current with AMCC - 4 Core','HorizontalAlignment'...
,'center','VerticalAlignment', 'top','FontWeight','bold',...
    'FontSize',14);

%=========================================================================
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
%=========================================================================



%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
%%Title: Electromagnetic Harvesting to power Energy Management Sensors%
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%

%This MATLAB code is used to determine the following:

% 1. Plot experimental measurements of induced Voltages across 470 ohms
% 2. Plot experimental measurements of current through 470 ohms load
```

```
%*********************************************************************
% Electromagnetic Harvesting to power Energy Management Sensors
clear all
close all
clc

% parameters for current carrying conductor
L = 15; % length (mm) copper #12 AWG conductor
C = 2.05/2; % conductor average radius (mm), NEC Table 8

%=====================================================================
Pl = [75; 150; 650; 725; 800; 1500; 1650; 2150; 2300];% Primary Load

% Induced voltage #25 AWG wire - 200 turns, AMCC-4 Core (connected load
=
% 470ohms) and short circuit current Iso (mA)

Vso_aa = [0.02; 0.06; 0.35; 0.39; 0.42; 0.83; 0.89; 1.18; 1.25];

Iso_aa = [1.54; 4.59; 24.95; 28.03; 29.69; 57.6; 62.3; 81.6; 87.2];

% Induced voltage #25 AWG wire - 300 turns, AMCC-4 Core (connected load
=
% 470ohms) and short circuit current Iso (mA)

Vso_bb = [0.06; 0.08; 0.53; 0.6; 0.63; 1.27; 1.37; 1.82; 1.92];

Iso_bb = [1.01; 3.02; 16.32; 17.34; 19.32; 37.80; 40.80; 53.60; 56.8];

% Induced voltage #25 AWG wire - 400 turns, AMCC-4 Core (connected load
=
% 470ohms) and short circuit current Iso (mA)

Vso_cc = [0.04; 0.12; 0.69; 0.78; 0.83; 1.68; 1.81; 2.40; 2.52];

Iso_cc = [1.49; 2.27; 12.37; 13.13; 14.62; 28.87; 31.15; 40.5; 43.1];

% Induced voltage #25 AWG wire - 500 turns, AMCC-4 Core (connected load
=
% 470ohms) and short circuit current Iso (mA)

Vso_dd = [0.05;0.14; 0.87; 0.97; 1.03; 2.05; 2.22; 2.94; 3.09];

Iso_dd = [1.19; 1.82; 9.87; 11.09; 11.73; 23.18; 25.02; 32.9; 34.8];

%Plots of Lab results
figure(1)
subplot(2,2,1)
plot(Pl,Vso_aa,'LineWidth',2)
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);
```

```
ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (200 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on

subplot(2,2,2)
plot(Pl,Iso_aa,'LineWidth',2)
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Sec. Load Current (mA)'},'FontWeight','bold','FontSize',12);

title({'Sec. Load Current Vrs Primary Load (200 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on
subplot(2,2,3)
plot(Pl,Vso_bb,'LineWidth',2)
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (300 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on
subplot(2,2,4)
plot(Pl,Iso_bb,'LineWidth',2)
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Sec. Load Current (mA)'},'FontWeight','bold','FontSize',12);

title({'Sec. Load Current Vrs Primary Load (300 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on

ha = axes('Position',[0 0 1 1],'Xlim',[0 1],'Ylim',[0
1],'Box','off','Visible','off','Units','normalized', 'clipping' ,
'off');

text(0.5, 1,'\bf Experimental Induced Voltage and Equivalent Current
with AMCC - 4 prototype Core (Rs = 470 ohms)',...
    'HorizontalAlignment','center','VerticalAlignment',...
    'top','FontWeight','bold','FontSize',14);

figure(2)
subplot(2,2,1)
plot(Pl,Vso_cc,'LineWidth',2)
set(gca,'XTick',200:200:2400)
```

```
xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (400 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on
subplot(2,2,2)
plot(Pl,Iso_cc,'LineWidth',2)
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Sec. Load Current (mA)'},'FontWeight','bold','FontSize',12);

title({'Sec. Load Current Vrs Primary Load (400 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on
subplot(2,2,3)
plot(Pl,Vso_dd,'LineWidth',2)
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

title({'Induced Voltage Vrs Primary Load (500 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on
subplot(2,2,4)
plot(Pl,Iso_dd,'LineWidth',2)
set(gca,'XTick',200:200:2400)
xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Sec. Load Current (mA)'},'FontWeight','bold','FontSize',12);

title({'Sec. Load Current Vrs Primary Load (500 turns, #25 AWG)'},...
    'FontWeight','bold','FontSize',12);
grid on


ha = axes('Position',[0 0 1 1],'Xlim',[0 1],'Ylim',[0
1],'Box','off','Visible','off','Units','normalized', 'clipping' ,
'off');

text(0.5, 1,'\bf Experimental Induced Voltage and Equivalent Current
with AMCC - 4 prototype Core (Rs = 470 ohms)',...
    'HorizontalAlignment','center','VerticalAlignment',...
    'top','FontWeight','bold','FontSize',14);

%======================================================================
%%%%%%%%%%%%%%%%%%%%%%%%%%%% End %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
%======================================================================
```

```
%**************************************************************************

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
%%Title: Electromagnetic Harvesting to power Energy Management Sensors%
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%

%This MATLAB code is used to determine the following:

%1. Compare simulated voltages with measured voltages from experiment
%**************************************************************************
clear all
close all
clc

% parameters for current carrying conductor
L = 15; % length (mm) copper #12 AWG conductor
C = 2.05/2; % conductor average radius (mm), NEC Table 8

%=========================================================================
% Core properties
%=================
lm = 12.2e-2; %(m), Magnetic path length (from core data), AMCC-4 Core
lg = 120e-6; %(m), Air gap between cut cores(range 25-150 micro meter)
mu_m = 3000; % (H/m), permeability of core material (Amorphous, 2605)
mur = mu_m/(1 + mu_m *(lg/(lm))); %relative permeability of core
mu0 = 4e-7 * pi; % Magnetic permeability of free space or vacuum
mu = mu0 * mur;% (H/m), Effective permeability of core


%=========================================================================
% Calculating Magnetic Strength around current carrying conductor
%=========================================================================
% Primary Load (Watts)
Rp = 100:100:2400;

%Primary Voltage (V rms)
Vp = 120;

Ip = Rp/Vp;

r = (C:.025:30);% radial distances from current carrying
conductor,0.025...
                % increments

B = zeros(1160,1);
for  p =1:24;
    for i = 1:1160;
        B(i,p) = (mu * Ip(p) * L * 1e-3)/(2*pi*r(i)*1e-3 *...
            sqrt(4*(r(i)*1e-3)^2 + (L*1e-3)^2));
    end
end

%=========================================================================
```

```
% Selecting magnetic field strength values at radial distances from...
% current carrying conductor (1, 2, 3, 4, 5, 6, 7, 8, 9, 10mm ).
%Current range for #12 AWG conductor is 1 to 20A.
%=========================================================================

r1 =r(41:1160)'; % radial distances
Bac = B(41:1160,1:24); % Corresponding magnetic field strength
% H = [r1 B(41:1160,1:24)]; % Matrix of radial distances and magnetic
flux

%=========================================================================
% Calcuting Induced voltage//Secondary turns//Output Voltage
%=========================================================================

% Core Parameters
%====================
Ac = 1.11;% Cross sectional area of core in cm.sq, AMCC-4 Core.

%Operating Parameters
%====================
Np = 1; % Number of turns of primary conductor ~1
f = 60; % output frequency in Hz
Kf = 4.44; % Waveform factor, sinewave.

% Magnetic field strength at (,6,7,8,9,10mm)from current carrying...
% conductor is used for calculation ~ approx distance from magnetic
path length(lm or MPL)
%=========================================================================
D = [Bac(201,:)' Bac(241,:)' Bac(281,:)' Bac(321,:)' Bac(361,:)'];

%=========================================================================
% Calculating induced voltages for various number of magnetic wire
windings around core. Ns = 50, 100, 150, 200, 250, 300, 350, 400, 450,
500.
%=========================================================================
Ns = (50:50:500);
Vs = zeros(24,5);
for w = 1:5,
    for g = 1:24,
        for m = 1:10,
            Vs (g,w,m) = Ac * Kf * D(g,w) * f * Ns(m)/(1e4);
        end
    end
end
Vs%;

%=========================================================================
%Comparing simulated and experimental voltages
%=========================================================================
Pl = [75; 150; 650; 725; 800; 1500; 1650; 2150; 2300];% Primary Load

% Induced voltage #25 AWG wire - 200 turns, AMCC-4 Core (Open circuit
% voltage, Vso, V rms) and short circuit current Iso (mA)

Vso_a = [0.03; 0.05; 0.34; 0.35; 0.40; 0.82; 0.88; 1.17; 1.23];
```

```
Iso_a = [3.02; 4.58; 24.88; 26.41; 29.64; 57.7; 62.2; 82.3; 86.9];

% Induced voltage #25 AWG wire - 300 turns, AMCC-4 Core (Open circuit
% voltage, Vso, V rms) and short circuit current Iso (mA)

Vso_b = [0.05; 0.08; 0.52; 0.56; 0.62; 1.27; 1.37; 1.82; 1.92];

Iso_b = [1.98; 3.02; 16.40; 17.36; 19.45; 37.80; 40.80; 53.90; 56.9];

% Induced voltage #25 AWG wire - 400 turns, AMCC-4 Core (Open circuit
% voltage, Vso, V rms) and short circuit current Iso (mA)

Vso_c = [0.07; 0.1; 0.7; 0.75; 0.84; 1.7; 1.83; 2.43; 2.55];

Iso_c = [1.49; 2.28; 12.42; 13.10; 14.67; 28.5; 30.8; 40.6; 42.9];

% Induced voltage #25 AWG wire - 500 turns, AMCC-4 Core (Open circuit
% voltage, Vso, V rms) and short circuit current Iso (mA)

Vso_d = [0.08;0.13; 0.89; 0.93; 1.05; 2.12; 2.28; 3.04; 3.19];

Iso_d = [1.20; 1.82; 10.09; 10.53; 11.80; 23.0; 24.7; 33.01; 34.5];


clf
figure (1)
jp = figure('Color',[1 1 1]);
subplot(2,2,1)
plot1=plot (Rp, Vs(:,:,4));
hold on
plot(Pl,Vso_a,'LineWidth',2,'Color',[0 0 0])

annotation('textarrow',[0.34125 0.394375],...
    [0.843387755102041 0.767857142857143],'TextEdgeColor','none',...
    'String',{'Experimental plot'},'FontWeight','bold','FontSize',12);
% grid on
set(plot1(1),'LineWidth',1.0,'Marker','o');
set(plot1(2),'LineWidth',1.0,'Marker','square')
set(plot1(3),'LineWidth',2.0,'LineStyle','--');
set(plot1(4),'LineWidth',1.0,'Marker','diamond');
set(plot1(5),'LineWidth',1.0,'Marker','^');
set(gca,'XTick',200:200:2400)

xlabel({'Primary Load (W)'},'FontWeight','bold','FontSize',12);

ylabel({'Induced Voltage (Vrms)'},'FontWeight','bold','FontSize',12);

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WIDEBAND PHASED ARRAY

&

RECTENNA DESIGN AND MODELING FOR

WIRELESS POWER TRANSMISSION

A Dissertation

by

JONATHAN NOEL HANSEN

Submitted to the Office of Graduate Studies of
Texas A&M University
in partial fulfillment of the requirements for the degree of

DOCTOR OF PHILOSOPHY

December 2011

Major Subject: Electrical Engineering

Wideband Phased Array &

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Approved by:

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Major Subject: Electrical Engineering

ABSTRACT

Wideband Phased Array &

Rectenna Design and Modeling for Wireless Power Transmission.

(December 2011)

Jonathan Noel Hansen, B.S., Texas A&M University

Chair of Advisory Committee: Dr. Kai Chang

Microstrip patch antennas are the most common type of printed antenna due to a myriad of advantages which encourage use in a wide range of applications such as: wireless communication, radar, satellites, remote sensing, and biomedicine. An initial design for a stacked-patch, broadband, dual-polarized, aperture-fed antenna is tested, and some adjustments are made to improve performance. The design goal is to obtain a 3 GHz bandwidth centered at 10 GHz for each polarization.

Once the single-element design is finalized, it is used in a 4x1 array configuration. An array increases the gain, and by utilizing variable phase-shifters to each element, the pattern can be electronically steered in a desired direction. The phase-shifters are controlled by a computer running LabVIEW so that the array's steering angle can be easily adjusted. The result of this new phased array design is a wide bandwidth system with dual-polarization which can be electronically steered.

Rectennas (rectifying antennas) are used in wireless power transmission (WPT) systems to collect microwave power and convert this power into useable DC power.

They find use in many areas such as space power transmission, RFID tags, wireless sensors, and recycling ambient microwave energy.

The ability to simulate rectenna designs will allow for an easier method of analysis and tuning without the time and expense of repetitive fabrication and measurement. The most difficult part of rectenna simulation is a good diode model, and since different diodes have dissimilar properties, a model must be specific to a particular diode. Therefore, a method of modeling an individual diode is the most critical part of rectenna simulation. A diode modeling method which is based on an equivalent circuit and compatible with harmonic balance simulation is developed and presented. The equivalent circuit parameters are determined from a series of S-parameter measurements, and the final model demonstrates S-parameters in agreement with the measured data.

An aperture-coupled, high-gain, single-patch rectenna is also designed and measured. This rectenna is modeled using the presented method, and the simulation shows good agreement with the measured results. This further validates the proposed modeling technique.

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Lastly, I would like to thank my family, especially my father, who has been a continuing source of guidance, support, and encouragement.

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1. INTRODUCTION

This dissertation is divided into two main topics. The first topic covers the design of a patch antenna and the subsequent development of this antenna into a phased array system. The second topic discusses rectenna theory, design, and modeling.

Section 2 discusses the design of a stacked-patch, broadband, dual-polarized, aperture-fed antenna. Background on patch antennas and the utilized design features are given, and reasons for all design decisions are addressed. Design goals, fabrication techniques, and measurement results which characterize the final design are also discussed.

Section 3 presents background information on array theory. Advantages of phased arrays are first discussed, followed by a brief mathematical description of how they operate. A few characteristics of phased arrays are addressed, and finally a discussion of design choices which will affect these characteristics is given.

Section 4 discusses the development of a phased array system which utilizes the single-element antenna designed in Section 2. A 4x1 array configuration is created which requires a power divider to be designed and fabricated to properly distribute power to each of the antenna elements. Phase-shifters must also be integrated into the system in order to control the phased array, and these phase-shifters are controlled by using LabVIEW to control a Data Acquisition (DAQ) device which is linked to a

---

This dissertation follows the style of *IEEE Transactions on Microwave Theory and Techniques*.

connector box and a logic-conversion circuit. The phased array is integrated into a useable system, calibrated, and measured for various settings. The results are then analyzed and discussed.

Section 5 gives a history of rectennas and their use in wireless power transmission systems. A definition for the term "rectenna" is first given, and some typical applications of rectennas are listed. A chronological history of rectenna development is then provided, starting with Hertz and Tesla in the 1800's and concluding with today's interest in a space solar-power satellite.

Section 6 provides a theoretical discussion of rectenna and wireless power transmission technology. The wireless power transmission system as a whole is first discussed, and the rectenna is then described in more detail. A rectenna is analogous to a half-wave rectifier, so a mathematical explanation of this is provided. An analytical model of rectenna efficiency is also derived and presented. The section concludes with a discussion on the effect of input power and diode selection.

Section 7 discusses rectenna simulation and its impact on rectenna design. A rectenna model is then proposed which will aid in simulation and simplify rectenna analysis and tuning. The model is based on a series of S-parameter measurements which allow the extraction of equivalent circuit model parameters. The final model is analyzed for usability and its match to the measurements on which it is based.

Section 8 proposes a new rectenna design and discusses its design, fabrication, and measurement. The new design is also modeled using the method proposed in

Section 7, and the simulation results are compared to the measured results to further validate the proposed rectenna model.

Section 9 concludes this dissertation with a summary, a discussion of accomplishments, and recommendations for future work.

The appendices provide LabVIEW and Matlab code developed to aid in the work presented here.

2. SINGLE-ELEMENT ANTENNA DEVELOPMENT

A. Background

The microstrip patch antenna is widely used because of its many advantages and is "often described as one of the most exciting developments in antenna and electromagnetic history [1]." These advantages include: lightweight, small size, conformability, easy fabrication, low cost, efficiency, and simple integration with microwave integrated circuits. However, there are some disadvantages which include: narrow impedance bandwidth, low gain, and the excitation of surface waves (decreases efficiency) [1]-[4]. A thorough history can be found in [1].

Radiation of the microstrip patch antenna is determined from the electromagnetic field distribution between the patch and the ground plane. The radiation can also be described in terms of the surface currents which exist on the patch [2]. Radiation arises from discontinuities at the patch edges, and resonance is created when the patch length is approximately a half-wavelength. However, some of the field extends outside the bounds of the physical patch. This is called the fringing effect, and it causes the patch to behave slightly larger than its physical dimensions. The length of the patch can be approximated as [5]

$$L \approx 0.49\lambda_d = 0.49\frac{\lambda_0}{\sqrt{\varepsilon_r}} \tag{2.1}$$

where λ_d is the wavelength in the dielectric, λ_0 is the free space wavelength, and ε_r is the relative permittivity of the dielectric.

Stacked patches can be used to increase the bandwidth. By stacking a parasitic patch (of different size) on top of the fed patch, two resonant frequencies will now exist as a result of the two different sized patches. This will effectively widen the bandwidth [6].

The most common methods of feeding early microstrip antennas were by means of a microstrip line contacting the patch edge or a coaxial probe extending through the ground plane and contacting the patch. However, microstrip feeds promote spurious radiation which limits bandwidth [4], and coaxial probes require long probes which increase spurious radiation from the probe, amplify surface waves, and increase feed inductance [2]. Microstrip feeds and coaxial probes also produce cross-polarized radiation due to inherent asymmetries which produce higher order modes [4]. Alternatively, an aperture-coupled feed electromagnetically couples from a microstrip line on the lower substrate to the patch through an aperture in the ground plane. This method isolates the feedline from the patch so that the feedline does not interfere with the radiating element (due to the shielding effect of the ground plane). This also has the benefit of increasing polarization purity. Coupling and bandwidth can be improved by adjusting the shape and size of the aperture, the feedline dimensions, and the stub length [2]. It has been shown that the best coupling is achieved when the aperture is in the shape of a bowtie [7].

Substrate choice is an important design decision. In general, substrates with a lower dielectric constant will give higher efficiency and a larger bandwidth. Also, as substrate thickness is increased, bandwidth tends to increase while efficiency decreases [8]. Therefore, tradeoffs must be made in selecting a suitable substrate.

Polarization of a wave is described by the direction of the electric field as time advances. If the electric field is oriented in a single direction then the wave is linearly polarized; if it rotates as time advances then the wave is circularly or elliptically polarized [3]-[4]. Dual linear polarization indicates that the wave is composed of two independent orthogonal linear polarizations. An antenna with dual-polarization allows two separate signals to be transmitted or received on the same antenna. Since both polarizations should operate at the same frequency bandwidth, both dimensions of the patch should be the same length. This has a negative effect on the bandwidth of the antenna since unequal dimensions of a patch can be used to increase bandwidth [9].

B. Initial Design

The design goals are to create a flat-panel, dual-polarized antenna which is centered at 10 GHz and has a 3 GHz bandwidth. A high cross-polarization isolation of greater than 20 dB is also desirable.

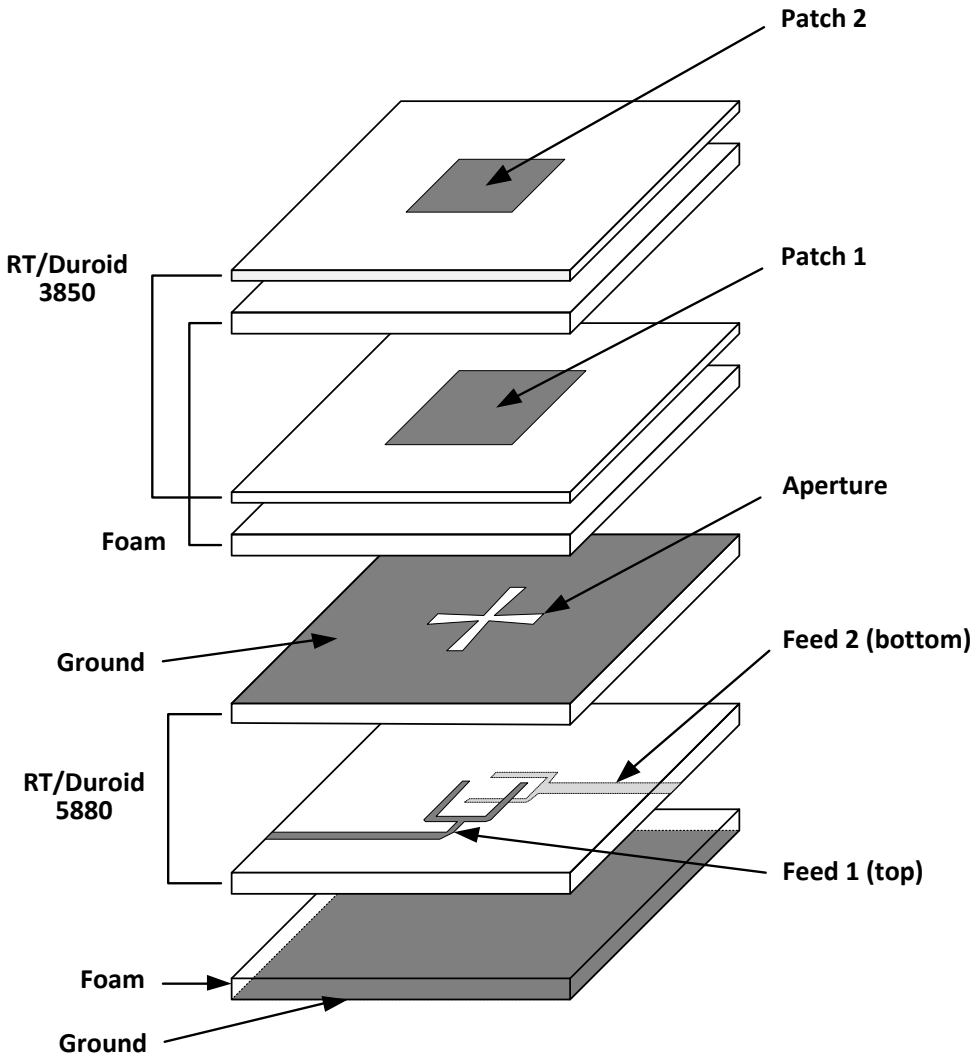


Figure 2.1. Single-element design (stacked-patch, aperture-fed, dual-polarized stripline).

The initial design of the single-element antenna [10] is a stacked-patch, aperture-fed, dual-polarized stripline design which is illustrated in Fig. 2.1. The patches are etched on 2 mil RT/Duroid 3850 substrate with a dielectric constant of 2.9. The top patch is supported by a 1.6 mm thick layer of lightweight foam, and the bottom patch is supported by a 3.2 mm thick layer of the same foam. This foam (Cuming C-FOAM PF-4/PSA) is specially made for microwave applications and has a low dielectric constant of 1.06. The apertured-ground plane and the feedlines are etched on 20 mil RT/Duroid 5880 substrate with a dielectric constant of 2.2. This places the top feed 20 mil from the apertured-ground plane and the bottom feed 40 mil from the apertured-ground plane.

There is also a second ground plane for the stripline design which is separated from the bottom feed by a 4.8 mm thick layer of the previously mentioned foam. The stripline scheme provides ground planes on both sides of the feedlines thereby shielding them from outside influences. The stripline structure also confines all of the electromagnetic field to the dielectric which allows it to support TEM waves and higher order TM/TE modes. This is in contrast to the microstrip design which has only one ground plane; this permits some of its field to exist in the air surrounding the dielectric which prevents the microstrip line from supporting pure TEM waves [11]. The downside to the stripline scheme is that since the field is distributed between two ground planes, the coupling through the aperture is reduced. However, by separating the second ground plane from the feedlines by a thick 4.8 mm layer of foam, most of the field is distributed toward the apertured-ground plane. This retains some of the benefit of a stripline feed while still maintaining good coupling through the aperture.

Dual-offset feedlines are used to improve impedance matching and decrease coupling between the feeds which reduces cross-polarization [9]. This feeding method also allows both feeds to share a single aperture. The feeding design can be seen in Fig. 2.2. It uses a simple two-way power divider with a mitered junction to split the power between the two feeds. The single input is a 50 Ω line which is split into two 100 Ω lines. Rather than extending straight to the edge of the structure as feed 2 does, feed 1 makes a 90° turn so that it exits the structure on the opposite side from feed 1. This is to ease its use in an array in which the feeds must exit on opposite sides.

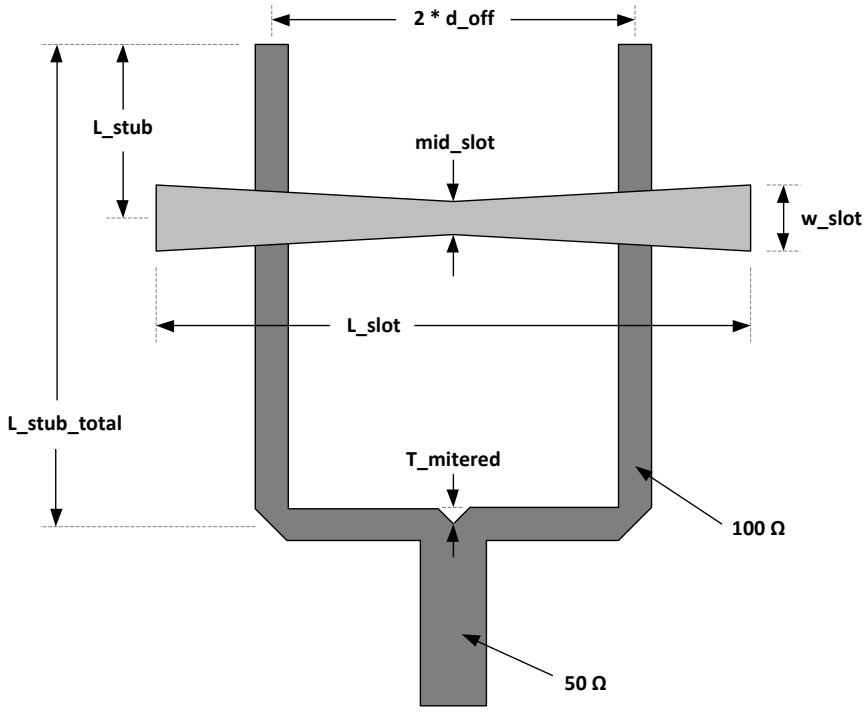


Figure 2.2. Dual-offset feedline with a bowtie aperture.

The design is tuned for a center frequency of 10 GHz and wide bandwidth. The dimensions of this design are given in Table 2.1.

Table 2.1. Initial antenna design parameters.

Variable	**Value**	**Variable**	**Value**
Height of feed 1 (top) substrate (h_feed1)	0.508 mm	Feed 1 (top) matching stub length (L_stub1)	1.98 mm
Height of feed 2 (bottom) substrate (h_feed2)	0.508 mm	Feed 2 (bottom) matching stub length (L_stub2)	2.07 mm
Height of patch 1 (bottom) foam (h_ant1)	3.2 mm	Feed 1 (top) total matching stub length (L_stub_total1)	8.6215 mm
Height of patch 2 (top) foam (h_ant2)	1.6 mm	Feed 2 (bottom) total matching stub length (L_stub_total2)	9.0165 mm
Feed 1 (top) width (100 Ω) (f1w_100)	0.403 mm	Aperture 1 length (L_slot1)	9.75 mm
Feed 1 (top) width (50 Ω) (f1w_50)	1.578 mm	Aperture 1 end-width (w_slot1)	1 mm
Feed 2 (bottom) width (100 Ω) (f2w_100)	1.013 mm	Aperture 1 mid-width (mid_slot1)	0.6 mm
Feed 2 (bottom) width (50 Ω) (f2w_50)	3.3 mm	Aperture 2 length (L_slot2)	10 mm
Offset feed 1 (top) separation (d_off1)	2.5 mm	Aperture 2 end-width (w_slot2)	1 mm
Offset feed 2 (bottom) separation (d_off2)	2.5 mm	Aperture 2 mid-width (mid_slot2)	0.6 mm
Mitered T-Junction Feed 1(top) (T1_mitered)	0.2015 mm	Patch 1 (bottom) length (L_p1)	11.3 mm
Mitered T-Junction Feed 2 (bottom) (T2_mitered)	0.5065 mm	Patch 2 (top) length (L_p2)	10.2 mm
Height of bottom foam (h_foam)	4.8 mm		

This design shows good simulation results using IE3D [12] as seen in Fig. 2.3. Using a standard VSWR cutoff value of 2.0, feed 1 shows a VSWR bandwidth of 42% (7.9 – 12.1 GHz) (f_0 = 10.0 GHz), and feed 2 shows a VSWR bandwidth of 44% (7.7 – 12.0 GHz) (f_0 = 9.85 GHz). The isolation between the two feeds is greater than 20 dB throughout the VSWR bandwidth.

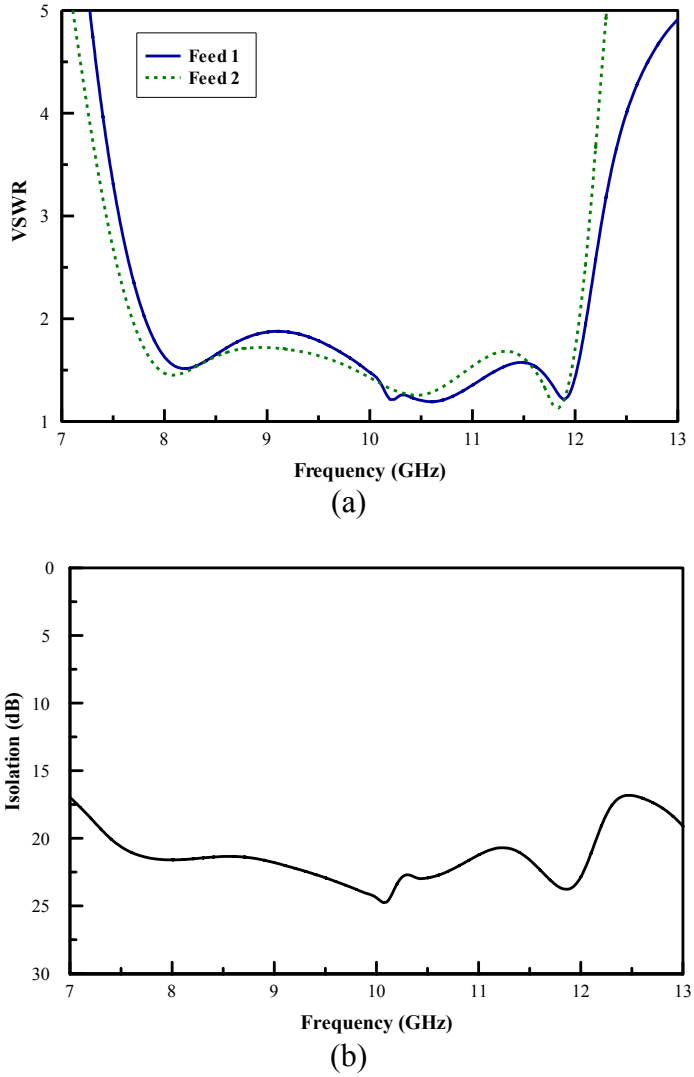


Figure 2.3. Simulation of the original single-element design. (a) VSWR and (b) isolation.

C. Fabrication of Initial Design

For optimal performance, the patches would be floating in the air or etched on top of the foam. Since this is not practical, a very thin substrate (2 mil) is placed on top of the foam layers. All patches, apertures, and feeds are etched using a photolithography process in which ultraviolet light and etching chemicals are used to remove specified areas of copper from the substrates.

All substrate and foam layers are cut as precisely as possible to the same dimensions as the ground planes (26 oz. zinc-plated steel is used as the bottom ground plane). The layers are then aligned and fixed together using 3M Super 77 Multipurpose Spray Adhesive. SMA connectors must also be soldered to the feedlines and both ground planes. Precise alignment is difficult, and care must be taken during this step since the alignment has a dramatic impact on performance of the antenna. A photograph of the fabricated antenna can be seen in Fig. 2.4.

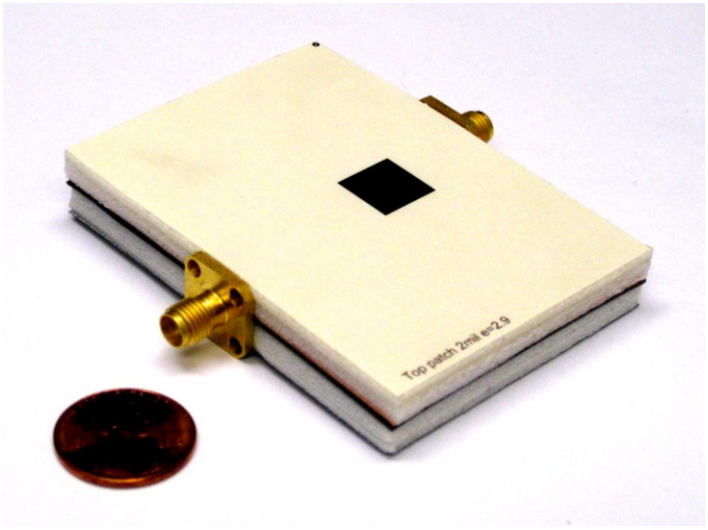


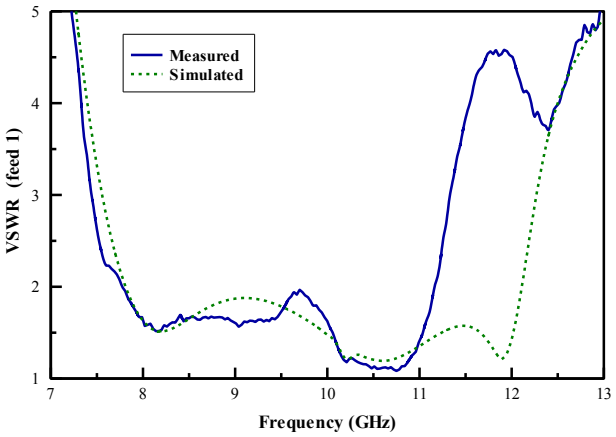
Figure 2.4. Photograph of the single-element antenna.

S-parameter measurements of the antenna can be seen in Fig. 2.5. Ignoring a small overrun in feed 2's VSWR, feed 1 has a measured bandwidth of 35% (7.8 – 11.1 GHz) (f_0 = 9.45 GHz), and feed 2 has a measured VSWR bandwidth of 37% (7.6 – 11.0 GHz) (f_0 = 9.3 GHz). The isolation between the two feeds is measured to be greater than 17 dB throughout the VSWR bandwidth. The measured results show a decrease in performance from the simulated values which is to be expected due to fabrication errors and extreme alignment difficulty. There is, however, one particularly troubling deviation from simulated data. The bandwidth's upper cutoff frequency is reduced by 1 GHz. For comparison, a number of devices were fabricated using the same procedure, and they all showed the same trend.

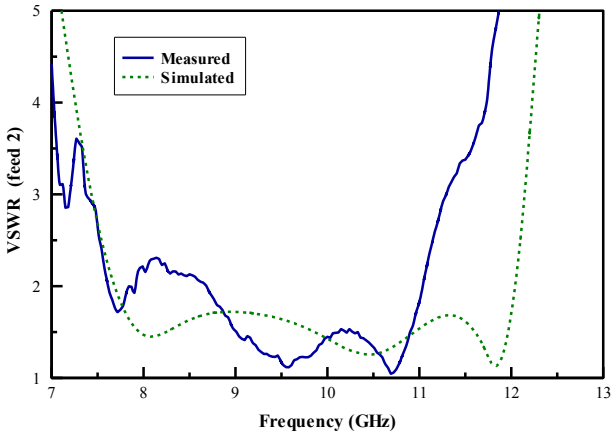
As a check on the IE3D [12] simulation results, another simulator, HFSS [13] was used to verify the solution. IE3D uses a method-of-moments analysis while HFSS uses a 3D finite element method. The simulated results from HFSS were very similar to the results from IE3D as can be seen in Fig. 2.6. Hence, comparable results were obtained from two different types of simulations, so simulator problems cannot explain the deviation in measured results.

As an attempt to increase alignment accuracy and possibly reduce any air gap errors in fabrication, a new fabrication method was tested in which feed 1 is etched on the bottom side of the apertured-ground plane substrate rather than on the top side of the feed 2 substrate. Several devices using this method were fabricated and compared to the previous method. This new method did show more consistent results, but the results were similar to the previous method, and the frequency-shifting problem still persisted.

Adhesive effects on multi-layer antenna performance have been analyzed in [14]-[15], and an explanation for the type of fabrication error appearing here is presented in the analysis. The authors theorize that the lossy adhesive used to bond the layers is the cause of the VSWR bandwidth's upper frequency shift. As an attempt to avoid adhesive losses, antennas were fabricated using double-sided tape to bond the layers instead of spray adhesive. The tape was strategically placed to avoid the areas where electromagnetic waves are coupling between layers. Measurements showed that this did not correct the problem, and the results were less consistent. This is probably due to air gaps introduced by the taping method.



(a)



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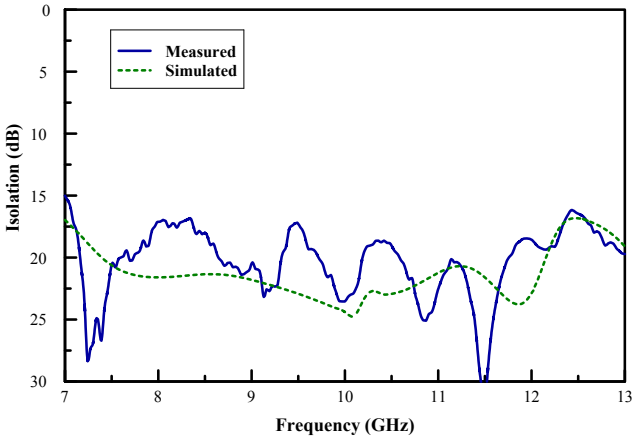
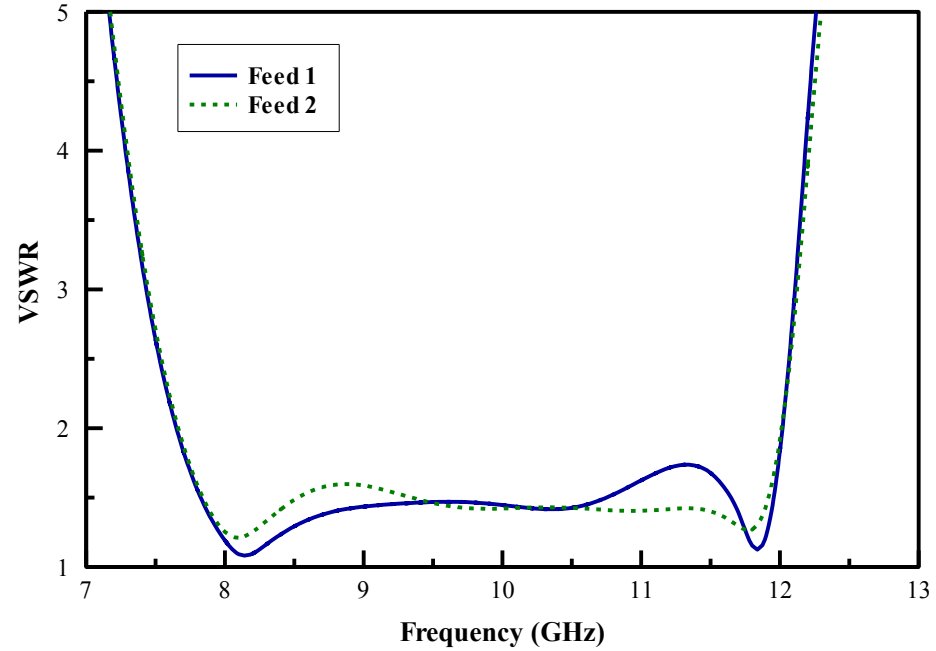
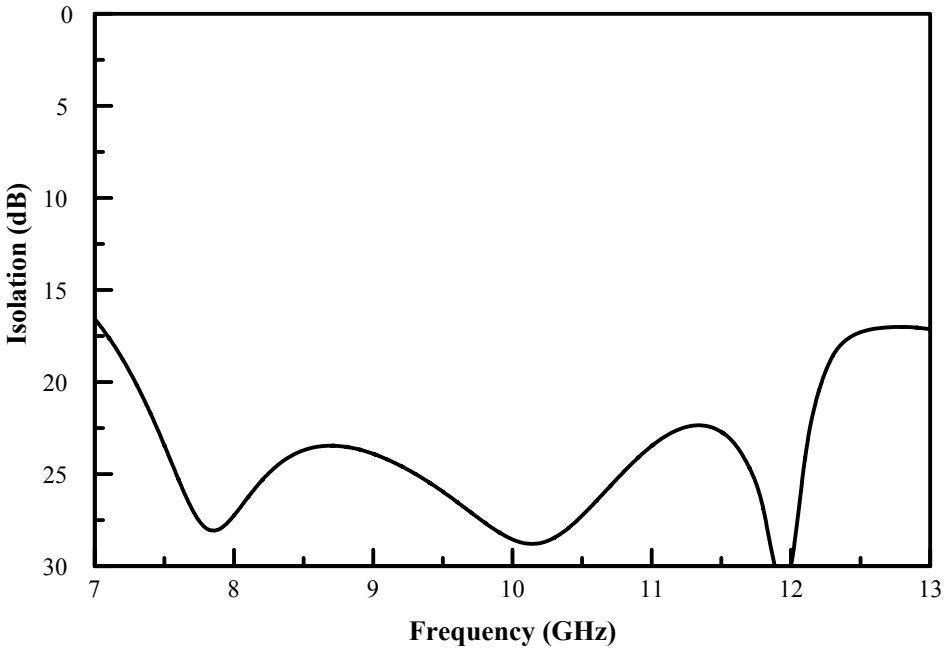


(c)

Figure 2.5. Measured and simulated VSWR and isolation of the original single-element design.



(a)



(b)

Figure 2.6. HFSS simulation of the original single-element design. (a) VSWR and (b) isolation.

D. Redesign (Dimension Adjustment) (Final Design)

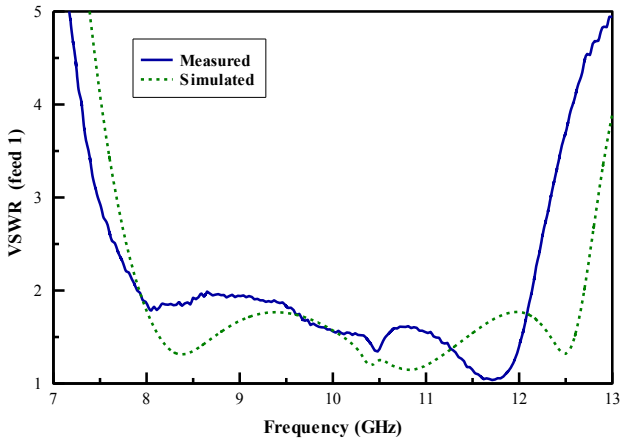
Since a fabrication method to correct the problem could not be found, the antenna was redesigned to give a higher upper frequency cutoff. The main change was a reduction in patch size which raised the resonant frequency. Some other parameters were then altered to fine tune the new design. The dimensions of the new design are given in Table 2.2.

Table 2.2. Final antenna design parameters.

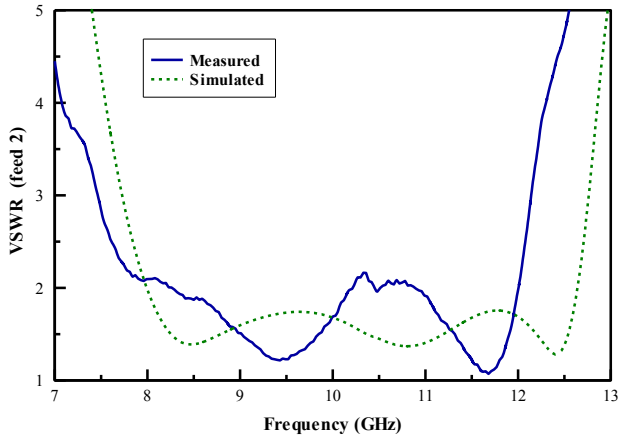
Variable	**Value**	**Variable**	**Value**
Height of feed 1 (top) substrate (h_feed1)	0.508 mm	Feed 1 (top) matching stub length (L_stub1)	1.96 mm
Height of feed 2 (bottom) substrate (h_feed2)	0.508 mm	Feed 2 (bottom) matching stub length (L_stub2)	2.07 mm
Height of patch 1 (bottom) foam (h_ant1)	3.2 mm	Feed 1 (top) total matching stub length (L_stub_total1)	8.6015 mm
Height of patch 2 (top) foam (h_ant2)	1.6 mm	Feed 2 (bottom) total matching stub length (L_stub_total2)	9.0165 mm
Feed 1 (top) width (100 Ω) (f1w_100)	0.403 mm	Aperture 1 length (L_slot1)	9.7 mm
Feed 1 (top) width (50 Ω) (f1w_50)	1.578 mm	Aperture 1 end-width (w_slot1)	1 mm
Feed 2 (bottom) width (100 Ω) (f2w_100)	1.013 mm	Aperture 1 mid-width (mid_slot1)	0.6 mm
Feed 2 (bottom) width (50 Ω) (f2w_50)	3.3 mm	Aperture 2 length (L_slot2)	9.6 mm
Offset feed 1 (top) separation (d_off1)	2.5 mm	Aperture 2 end-width (w_slot2)	1 mm
Offset feed 2 (bottom) separation (d_off2)	2.5 mm	Aperture 2 mid-width (mid_slot2)	0.6 mm
Mitered T-Junction Feed 1(top) (T1_mitered)	0.2015 mm	Patch 1 (bottom) length (L_p1)	10.9 mm
Mitered T-Junction Feed 2 (bottom) (T2_mitered)	0.5065 mm	Patch 2 (top) length (L_p2)	9.5 mm
Height of bottom foam (h_foam)	4.8 mm		

S-parameter simulation results and measurements of the new antenna design can be seen in Fig. 2.7. For the simulation, using a standard VSWR cutoff value of 2.0, feed 1 shows a VSWR bandwidth of 47% (7.9 – 12.7 GHz) (f_0 = 10.3 GHz), and feed 2 shows a VSWR bandwidth of 45% (8.0 – 12.6 GHz) (f_0 = 10.3 GHz). The isolation between the two feeds is greater than 20 dB throughout the VSWR bandwidth.

Ignoring a small overrun in feed 2's VSWR, feed 1 has a measured bandwidth of 42% (7.9 – 12.1 GHz) (f_0 = 10.0 GHz), and feed 2 has a measured VSWR bandwidth of 38% (8.2 – 12.0 GHz) (f_0 = 10.1 GHz). The isolation between the two feeds is measured to be greater than 17 dB throughout the VSWR bandwidth. The measured results still show a shift in the bandwidth's upper cutoff frequency, but now that shift is accounted for, so the measured center frequencies for feeds 1 and 2 are 10.0 GHz and 10.1 GHz, respectively. This dimension adjustment has corrected the fabricated antenna and shifted the bandwidth to the desired location.



(a)



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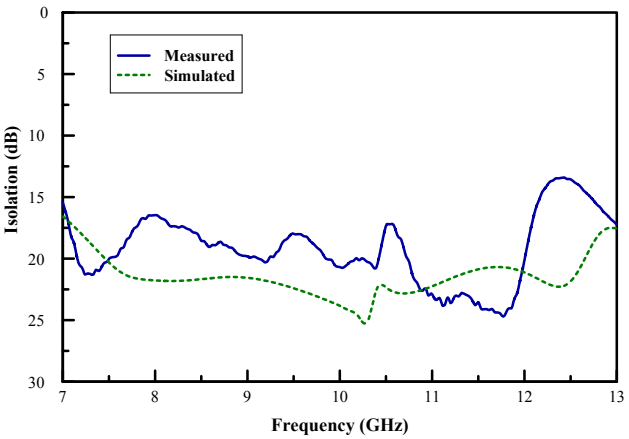


(c)

Figure 2.7. Measured and simulated VSWR and isolation of the redesigned (final) single-element with new dimensions.

E. Radiation Pattern Measurement

The radiation pattern of the redesigned (final) antenna was measured in an anechoic chamber. Each feed is measured separately, and when one feed is being measured, the other is terminated with a 50 Ω load to prevent reflections. Also, during measurement it is important to cover the antenna's connectors with absorber to prevent reflections. If this is not done, results similar to Fig. 2.8 will be obtained.

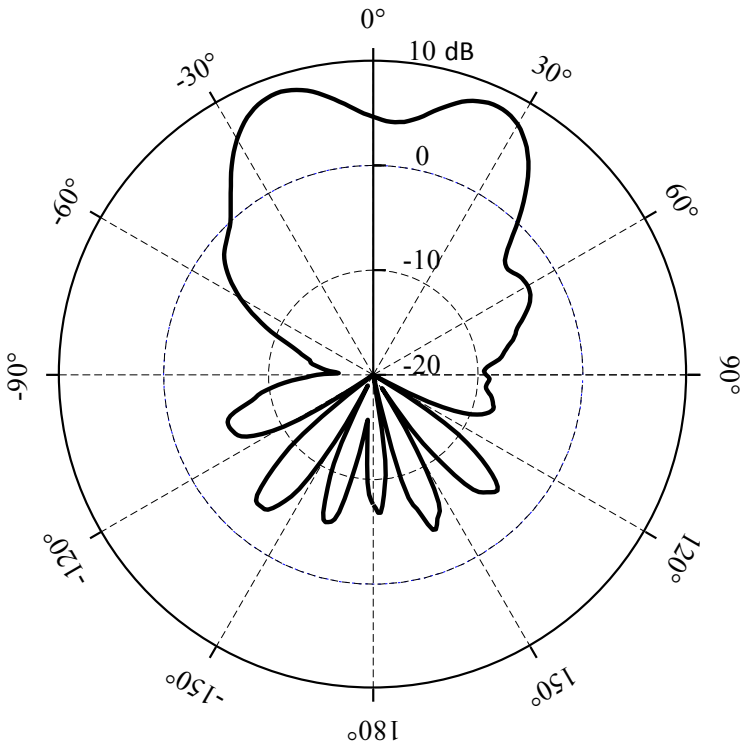


Figure 2.8. Radiation pattern with no absorber (10 GHz / Port 1 / co-pol E-plane).

After a complete set of antenna measurements was taken, a comparison was made to simulated results (Fig. 2.9), and it was noticed that the absorber covering the connectors might be disturbing some of the measured pattern at larger measurement angles and artificially reducing the beamwidth of the antenna. This was corrected by fabricating a new antenna with a larger ground plane (Fig. 2.10). This placed the connectors farther away from the patch so that when they are covered with absorber, it does not interfere with the pattern measurement. This adds approximately 15-25° of beamwidth to the measured pattern. This is illustrated in Fig. 2.11.

Measured radiation patterns of the antenna are displayed in Fig. 2.12 for 8.5, 10, and 11.5 GHz. The patterns are well behaved, and the cross-polarization isolation is greater than 20 dB for all cases. Also, graphs of gain and half-power beamwidth (HPBW) versus frequency are displayed in Figs. 2.13-2.14. The measured gain ranges from 5 - 8 dBi, and the measured HPBW is 50 - 70° for the E-plane and 70 - 90° for the H-plane. Accounting for measurement error and alignment difficulties during fabrication, these results match well to the simulation.

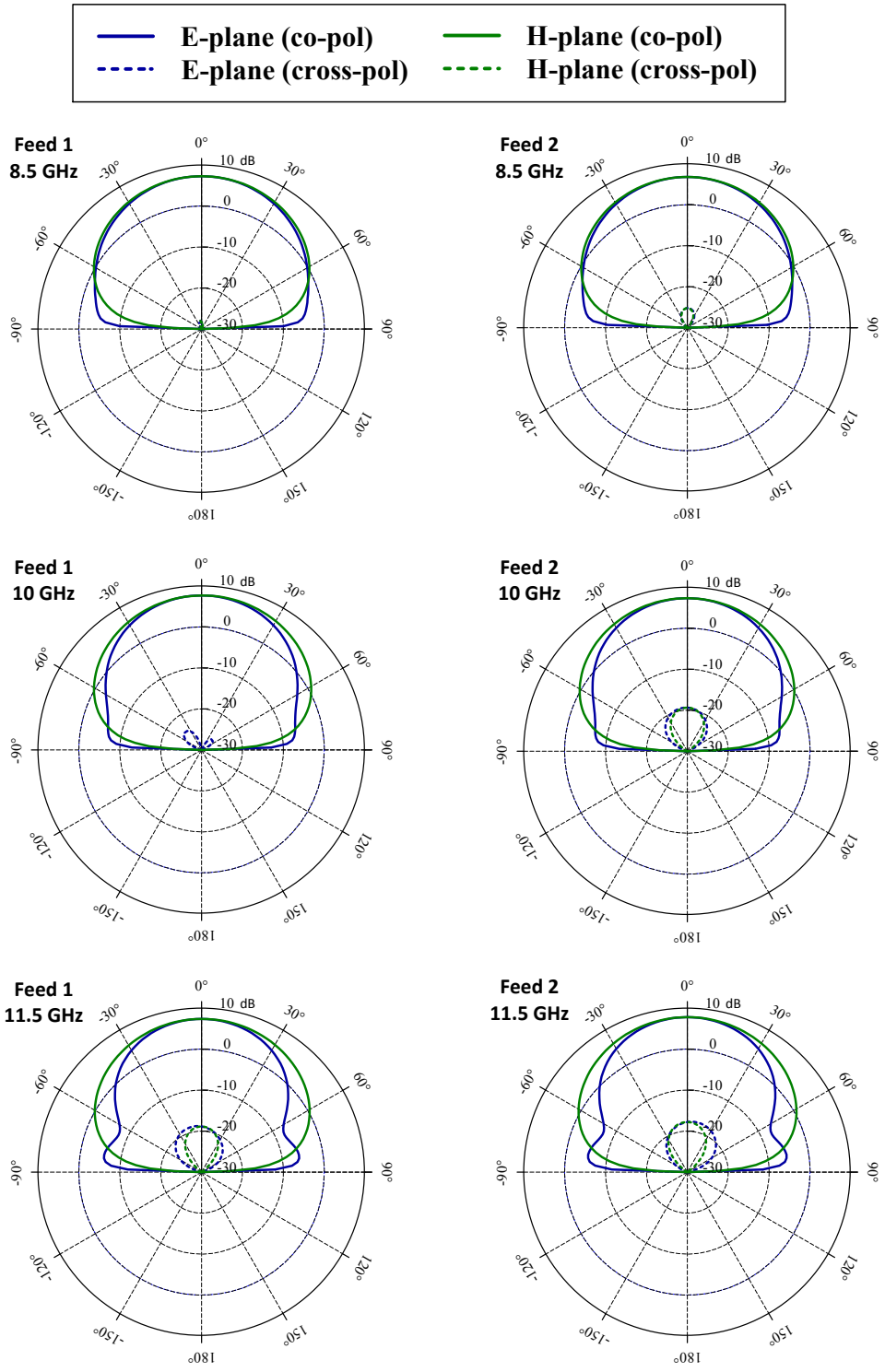


Figure 2.9. Simulated radiation pattern of the single-element final design.

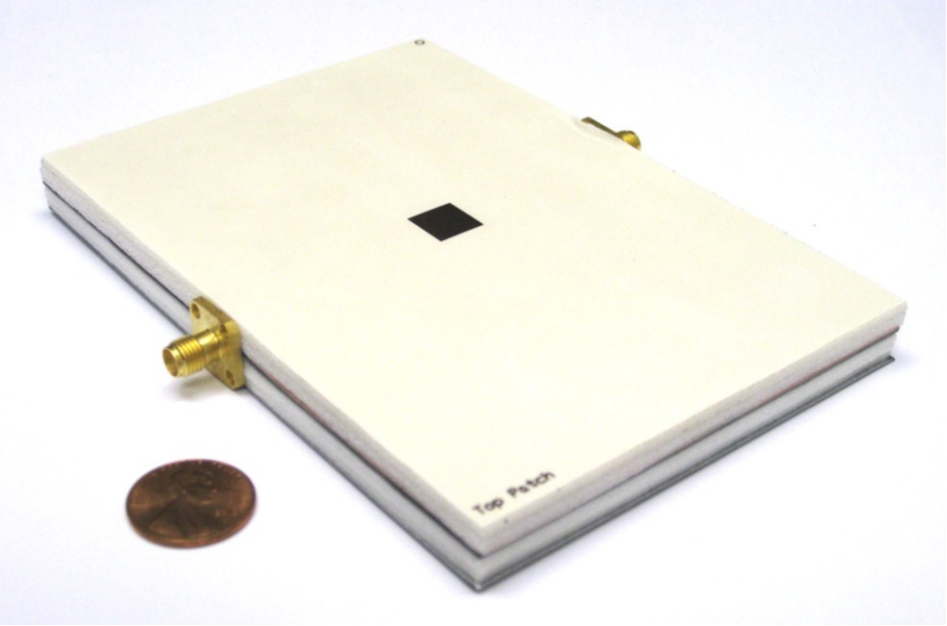


Figure 2.10. Photograph of the single-element antenna with larger ground plane.

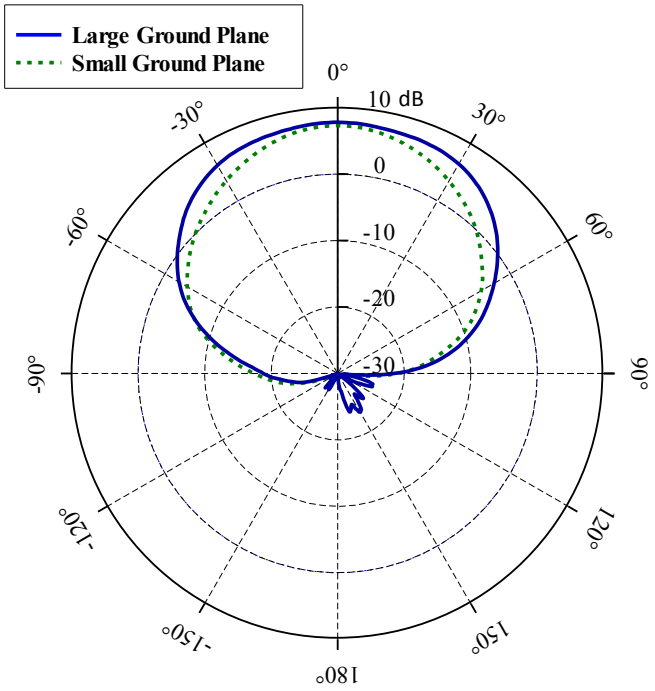


Figure 2.11. Measured radiation pattern comparison of ground plane size (10 GHz / Port 2 / co-pol H-plane).

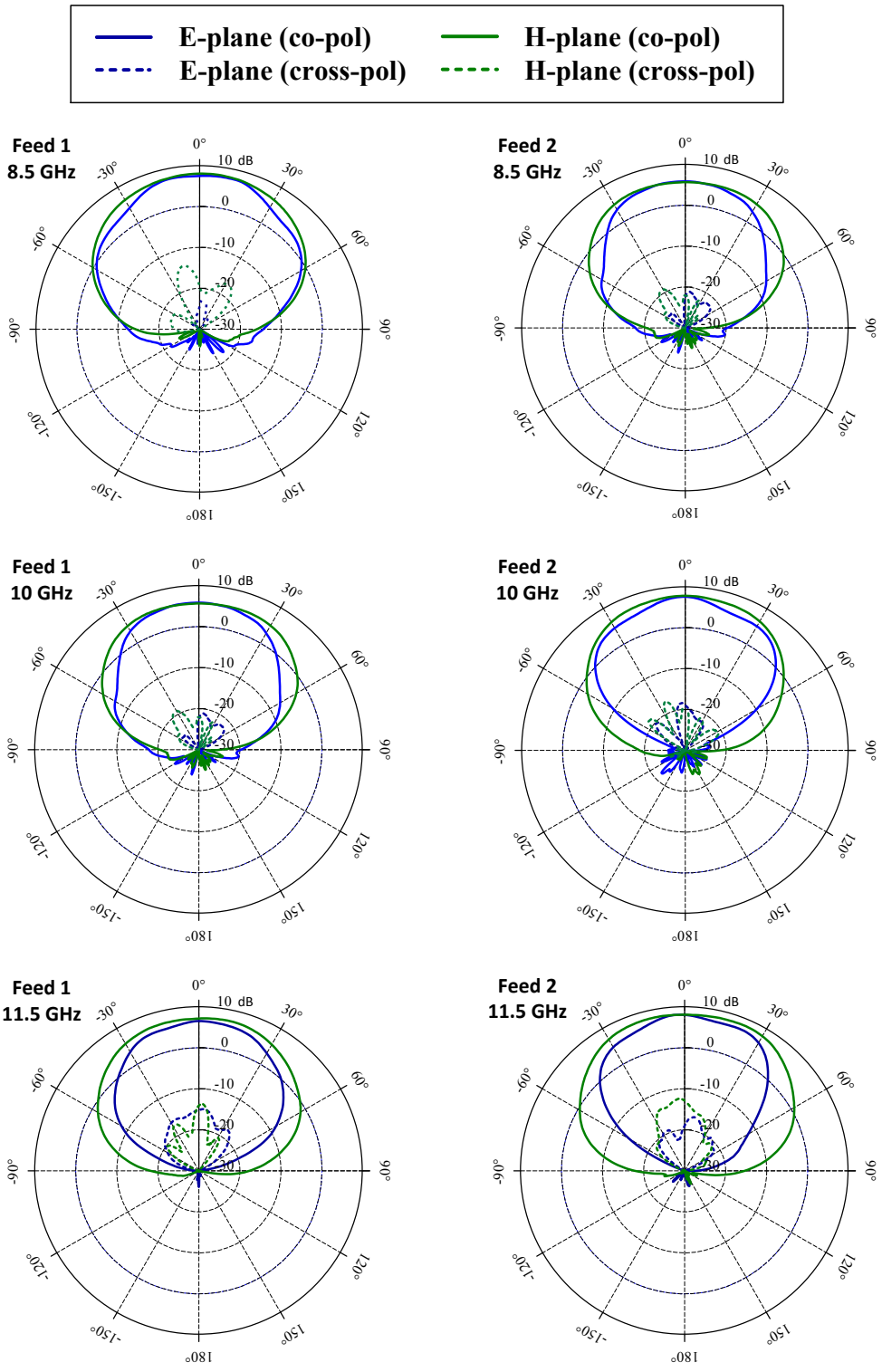


Figure 2.12. Measured radiation pattern of the single-element final design.

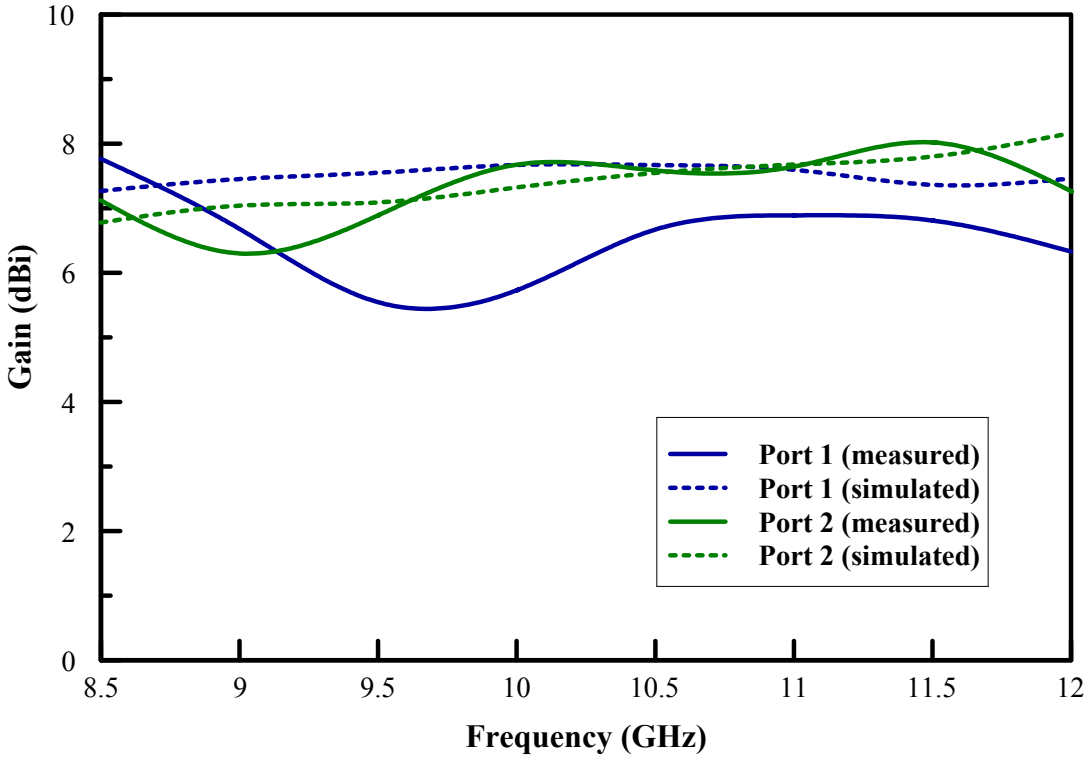
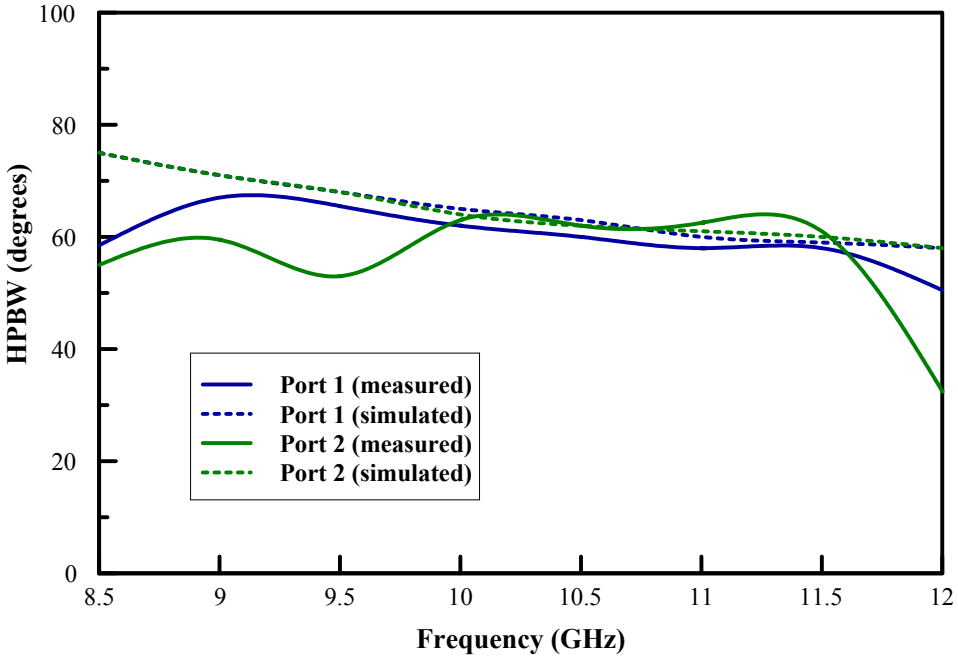
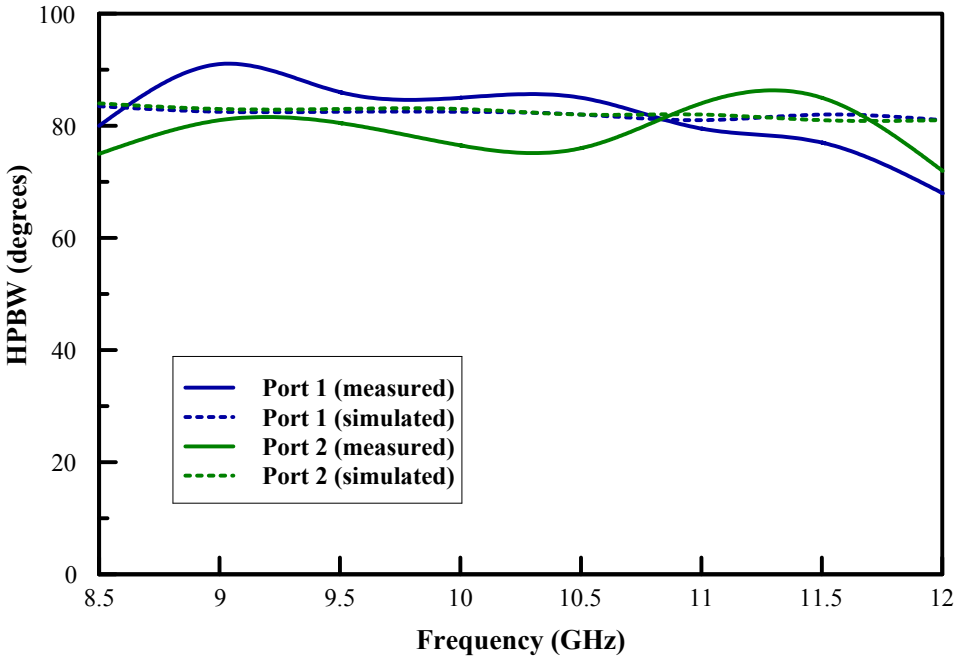


Figure 2.13. Gain versus frequency of the radiation pattern for the single-element final design.



(a)



(b)

Figure 2.14. Half-power beamwidth (HPBW) versus frequency of the radiation pattern for the single-element final design. (a) E-plane and (b) H-plane.

F. Conclusion

S-parameter measurements of the initial single-element design showed a shift in the VSWR bandwidth's upper frequency cutoff which shifted the bandwidth out of the desired range of 8.5 – 11.5 GHz. Some adjustments to the antenna dimensions corrected for this unavoidable shift, and the new design was shown to perform well in the target bandwidth and beyond. The VSWR bandwidth of feed 1 was measured as 42% with f_0=10.0 GHz, and the VSWR bandwidth of feed 2 was measured as 38% with f_0=10.1 GHz. S-parameter isolation between the feeds was measured to be greater than 17 dB, and gain measurements showed a cross-polarization isolation of greater than 20 dB. Gain was measured to be 5 – 8 dBi throughout the target bandwidth, and HPBW was measured to be 50 – 90°. Measurements and simulation of the antenna show a good match, and the antenna should work well in a phased array application.

3. ARRAY THEORY

A. Introduction

Single-element antennas have two significant drawbacks: a large beamwidth and a fixed beam direction. This can be overcome with an array of antenna elements which is able to focus energy in a desired direction to narrow the beam and increase range. In order to increase the directivity, the electrical size of an antenna must be increased, and creating an array of elements is a simple and effective way of doing this [4]. An antenna can be mechanically steered, but this is slow and prone to breakage [16]. Alternatively, an antenna array can be fitted with phase shifting devices which will allow the real-time adjustment of individual elements so that the elements will interfere constructively in the desired direction and interfere destructively in all other directions.

Phased arrays were first used in the 1930s for communications between the United States and England. It was, however, the radar and radioastronomy communities that were responsible for the large advances in phased array technology [17]. Since a phased array radar steers without movement and inertia, it has the ability to pause scans to check detections, track multiple targets by switching back and forth, and simultaneously survey and track [16].

B. Array Factor

An *N*-element linear array can be formed by aligning *N* identical antenna elements along the z-axis with each element separated by a distance *d* and phase β. This *N*-element array will form a plane wave whose direction is determined by the progressive phase shift between elements [3]. This is illustrated in Fig. 3.1 where each element has a phase-shifter with a shifting value of β degrees greater than the previous element. This phase shift offset will align the phases to form a wave front and create a plane wave traveling at an angle θ_0 with respect to the array's broadside. The angle θ is with respect to the z-axis on which the elements are positioned.

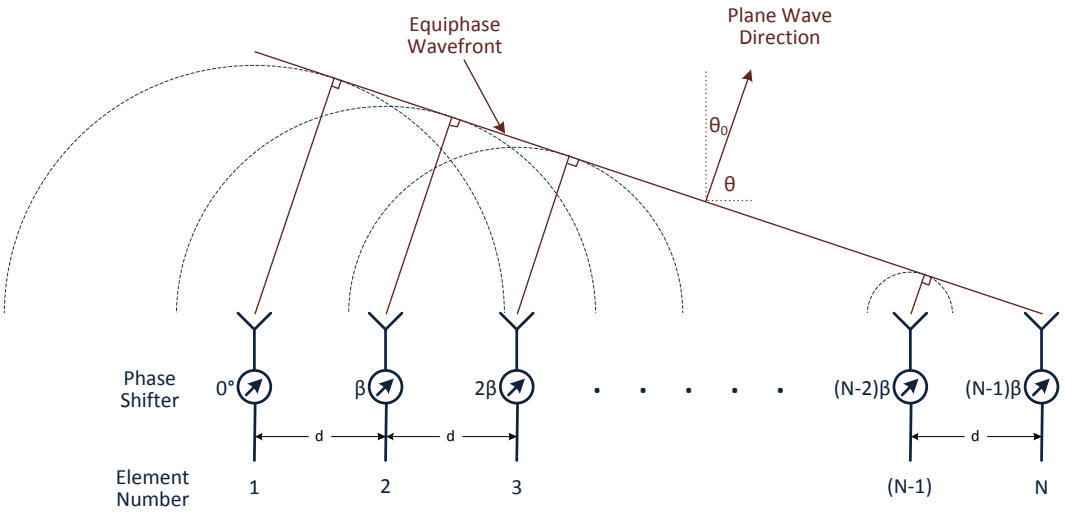


Figure 3.1. An *N*-element linear array with each element separated by a distance *d* and phase β.

The total radiated field from an array of identical single-elements can be calculated as the sum of all the individual elements. The far-field approximation [4][3]:

$$\theta_1 = \theta_2 = \ldots = \theta_N = \theta \tag{3.1}$$

$$\left.\begin{aligned} r_1 &= r \\ r_2 &= r + d\cos\theta \\ &\vdots \\ r_N &= r + (N-1)d\cos\theta \end{aligned}\right\} \text{ for phase variation} \tag{3.2}$$

$$r_1 = r_2 = \ldots = r_N = r \qquad \text{for amplitude variation} \tag{3.3}$$

can be applied in order to express the total field from the array as

$$E_{\text{total}} = E_{\text{single-element}} \times AF \tag{3.4}$$

where $E_{\text{single-element}}$ is the electric field of the single-element at the reference point and AF is the array factor. For an array of uniform amplitude and spacing, the array factor can be expressed as [4]

$$AF = \sum_{n=1}^{N} e^{j(n-1)\psi} \tag{3.5}$$

$$\text{where} \qquad \psi = kd\cos\theta + \beta \ . \tag{3.6}$$

The parameter *k* is the wave number, *d* is the distance between elements, and β is the progressive phase shift between elements. By setting the reference point to the center of the array, applying complex exponential identities, and normalizing [4], the array factor can also be expressed as

$$AF = \frac{1}{N}\left[\frac{\sin\left(\frac{N}{2}\psi\right)}{\sin\left(\frac{1}{2}\psi\right)}\right] \quad . \tag{3.7}$$

As described in Fig. 3.1, the progressive phase shift β produces a wave front moving in the direction of the angle θ_0 (where $\theta_0 = 90° - \theta$). The array can then be steered by adjusting β. As seen in (3.5), the array factor (*AF)* is maximum when ψ=0, so the array's scan angle can be calculated by

$$\beta = -kd\cos\theta = -kd\sin\theta_0 . \tag{3.8}$$

For an N-element array with elements separated by half-wavelengths, the half-power (3 dB) beamwidth can be estimated [3] as

$$\theta_{HPBW} \approx \frac{100}{\sqrt{N}} \tag{3.9}$$

where θ_{HPBW} is in degrees. It should also be mentioned that the AF's beamwidth increases some with increasing scan angles.

C. Grating Lobes

Since the array factor is a periodic function, it is possible for constant phase wave fronts to form by constructive interference in more than one direction. The additional wave fronts which appear are known as grating lobes which occur when the array factor's exponential in (3.5) is equal to a multiple of 2π, so grating lobes appear when

$$\psi = \pm 2\pi m \quad , \quad \text{m} = 0, 1, 2, \ldots \tag{3.10}$$

Then to scan to an angle θ_0, the condition in (3.8) must hold, so (3.6) and (3.10) can be equated and (3.8) substituted for β resulting in the condition

$$\psi = kd\cos\theta - kd\sin\theta_0 = \pm 2\pi m \quad . \tag{3.11}$$

Next, the observation angle θ (relative to the z-axis) will be changed to an angle relative to broadside and identified as the grating lobe angle such that $\theta_g = 90° - \theta$. Then the condition for grating lobes becomes

$$kd(\sin\theta_g - \sin\theta_0) = \pm 2\pi m \quad , \tag{3.12}$$

and since $k=2\pi/\lambda$,

$$\frac{d}{\lambda}\left(\sin\theta_g - \sin\theta_0\right) = \pm m \ . \tag{3.13}$$

Then the grating lobe angles can be found from

$$\sin\theta_g = \pm m\frac{\lambda}{d} + \sin\theta_0 \quad , \tag{3.14}$$

and the condition for avoiding grating lobes in all of real space is

$$\left|\sin\theta_0 \pm \frac{\lambda}{d}\right| > 1 \tag{3.15}$$

which is satisfied if

$$\frac{\lambda}{d} > 1 + \sin\theta_0 \tag{3.16}$$

which can be rewritten as

$$\frac{d}{\lambda} < \frac{1}{1+\sin\theta_0} \quad . \tag{3.17}$$

Since the maximum scan is 90°, the condition to avoid grating lobes (in all visible space) for all scan angles is

$$d < \lambda/2 \quad . \tag{3.18}$$

It should be noted that the θ in (3.8) and the θ in (3.11) represent observation angles for different pattern features. The θ in (3.8) denotes the scan angle, and the θ in (3.11) denotes grating lobe angles. Both θs are with respect to the z-axis.

D. Non-Uniform Amplitude

Non-uniform amplitude distributions can be used to adjust the beamwidth and sidelobe levels of a phased array. The equation for the array factor in (3.5) can be modified to include the effect of element amplitudes a_n as

$$AF = \sum_{n=1}^{N} a_n e^{j(n-1)\psi} \quad . \tag{3.19}$$

By setting the reference point to the center of the array, applying complex exponential identities, and normalizing [4], the array factor can also be expressed as

$$(AF)_{2M}(\text{even}) = \sum_{n=1}^{M} a_n \cos\left[(2n-1)u\right] \tag{3.20}$$

$$(AF)_{2M+1}(\text{odd}) = \sum_{n=1}^{M+1} a_n \cos\left[2(n-1)u\right] \tag{3.21}$$

$$\text{where} \qquad u = \frac{\pi d}{\lambda}\cos\theta \ . \tag{3.22}$$

In these equations, the total number of elements for an even array is $2M$ and for an odd array is $2M+1$. Also, the amplitude excitation of the center element for an odd numbered array as described in (3.21) is $2a_1$.

Amplitude coefficients for each element can then be chosen and used in (3.19)-(3.22) to determine the array factor. Two of the most popular amplitude distributions are discussed here, but there are many others.

Binomial Distribution

A binomial distribution [4] can be found by expanding the function $(1\text{-}x)^{m-1}$ using a binomial expansion as

$$(1+x)^{m-1} = 1 + (m-1)x + \frac{(m-1)(m-2)}{2!}x^2 + \frac{(m-1)(m-2)(m-3)}{3!}x^3 + \cdots \qquad (3.23)$$

The positive coefficients of the series can then be used as amplitude coefficients for an array of *m* elements. These coefficients for increasing values of *m* form a Pascal triangle as:

m = 1							1							
m = 2						1		1						
m = 3					1		2		1					
m = 4				1		3		3		1				
m = 5			1		4		6		4		1			
m = 6		1		5		10		10		5		1		
m = 7	1		6		15		20		15		6		1	
m = 8	1		7		21		35		35		21		7	1

(3.24)

Dolph-Tschebyscheff Distribution

The Dolph-Tschebyscheff distribution [4] is formed from the Tschebyscheff polynomials and is a compromise between a uniform and a binomial distribution. Its amplitude coefficients can be chosen to create sidelobes of any desired level, and when a Dolph-Tschebyscheff array with no sidelobes is chosen, it reduces to a binomial array. The Dolph-Tschebyscheff polynomial [18] is defined as

$$T_m(z) = \cos\left[m\cos^{-1}(z)\right], \qquad -1 \le z \le 1 \tag{3.25}$$

$$T_m(z) = \cosh\left[m\cosh^{-1}(z)\right], \qquad |z| > 1 \quad . \tag{3.26}$$

The method of finding the Dolph-Tschebyscheff coefficients involves expanding (3.25)-(3.26) and the array factor to be of the same form and equating the coefficients. The complete procedure is explained in detail in [4] and [18]. The Dolph-Tschebyscheff coefficients will give an array with all sidelobes at a level specified in the design procedure. However, decreasing the sidelobe level increases the beamwidth and the difference in coefficient magnitudes.

E. Phase Quantization

Most recent phased array systems use digital phase-shifters to control the phase of each element. This effectively quantizes the phase so that the realized phase is only an approximation of the ideal phase. For an *N*-bit phase-shifter, the phase states are separated by the least significant phase ϕ_0 which is given by

$$\phi_0 = \frac{2\pi}{2^N} \quad . \tag{3.27}$$

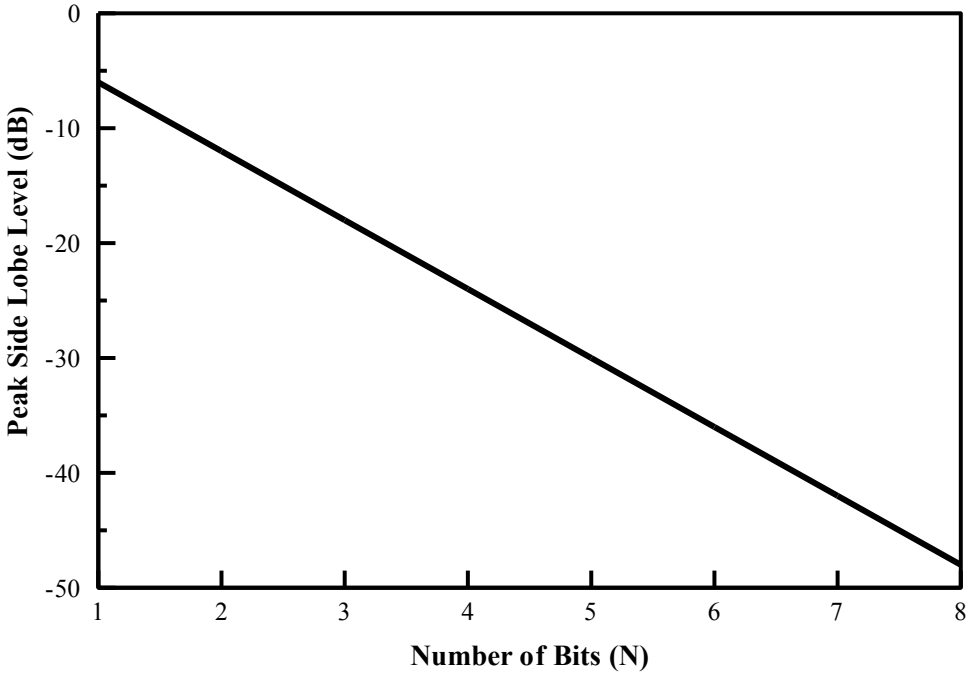
This phase approximation creates phase quantization grating lobes, and the first detailed analysis of this effect was performed by Miller [19]-[20]. He evaluated the level of the first grating lobe by assuming that the array current distribution is continuous (i.e., there is not a discrete number of elements). From this analysis, the level of the first phase quantization lobe can be expressed as

$$P_{QL} = \frac{1}{2^{2N}} \approx -6N \text{ (dB)} . \tag{3.28}$$

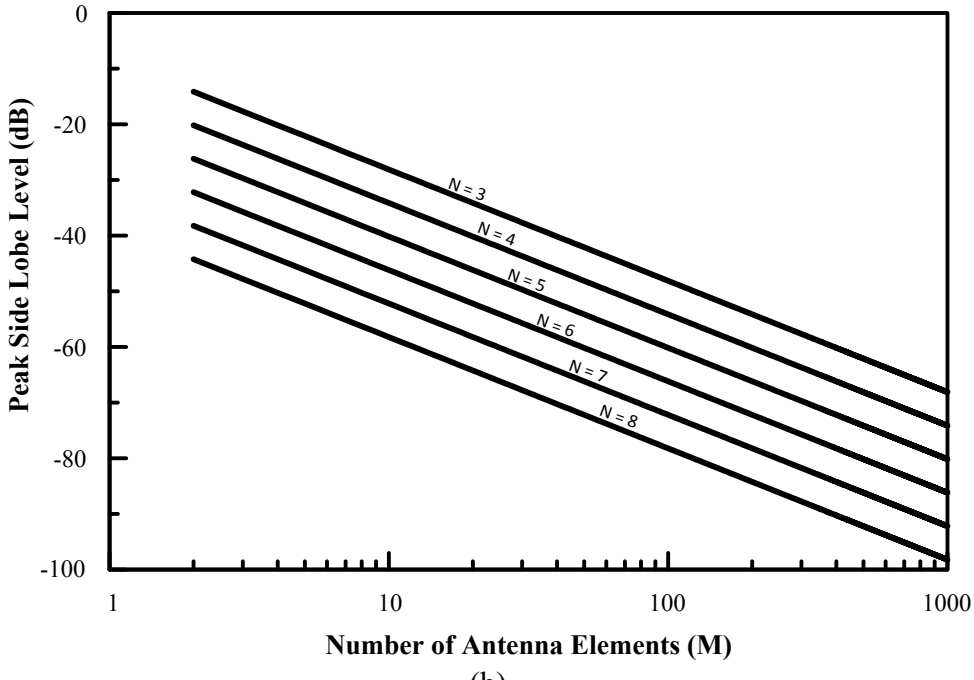
The continuous array approximation used by Miller led to an underestimation of the actual peak phase quantization lobes [21]. More accurate equations have been given by various authors including Hansen [22] and Mailloux [23]. Mailloux gives an estimate for the upper bound on the phase quantization sidelobe level as

$$P_{QL} < -20\log M + 9.94 - 6.02N \text{ (dB)} \quad \text{for } M \geq 2 \text{ and } N \geq 3 \tag{3.29}$$

where N is still the number of bits and M is the number of array elements. The equations for phase quantization sidelobe levels as proposed by Miller (3.28) and Mailloux (3.29) are plotted in Fig. 3.2.



(a)



(b)

Figure 3.2. Maximum phase quantization sidelobe levels as proposed by (a) Miller and (b) Mailloux.

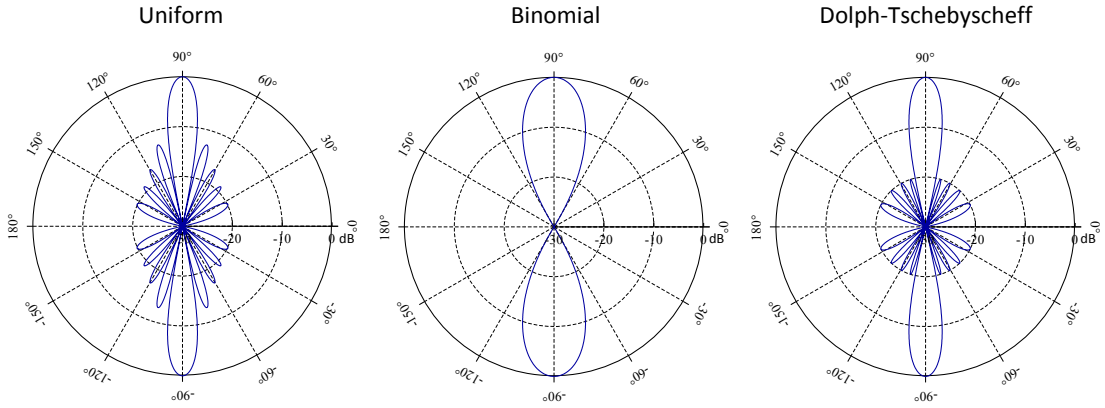
F. Design Choices

When designing a phased array, decisions must be made which balance the tradeoffs between performance and cost/complexity. Assuming that the single-element is already designed, these decisions involve the amplitude distribution, the number of elements, element spacing, and the number of phase-shifter bits.

Amplitude Distribution

As previously mentioned, there are many choices for an array's amplitude distribution, of which the most popular are: uniform, binomial, and Dolph-Tschebyscheff. Examples of these are illustrated in Fig. 3.3. One of the uniform distribution's advantages is that it has a beamwidth which is narrower than the other distributions. The downside is that the sidelobes are relatively large. The uniform distribution is also simple to implement; since all of the elements' amplitudes are equal, there is no need for an amplifier or attenuator for each feed. For non-uniform distributions, individual element amplitudes must be controlled, which leads to increased system complexity and cost.

The binomial distribution eliminates sidelobes, but it does this at the expense of a wide beamwidth. Also, the difference in element amplitudes is quite large, and this can be difficult to implement in a practical system.



N = 10 d = λ/2

	Uniform	Binomial	Dolph-Tschebyscheff
	1	1	1
	1	9	1.67
	1	36	2.60
	1	84	3.41
Amplitude	1	126	3.88
Distribution	1	126	3.88
	1	84	3.41
	1	36	2.60
	1	9	1.67
	1	1	1
HPBW	10.2°	20.2°	11.2°

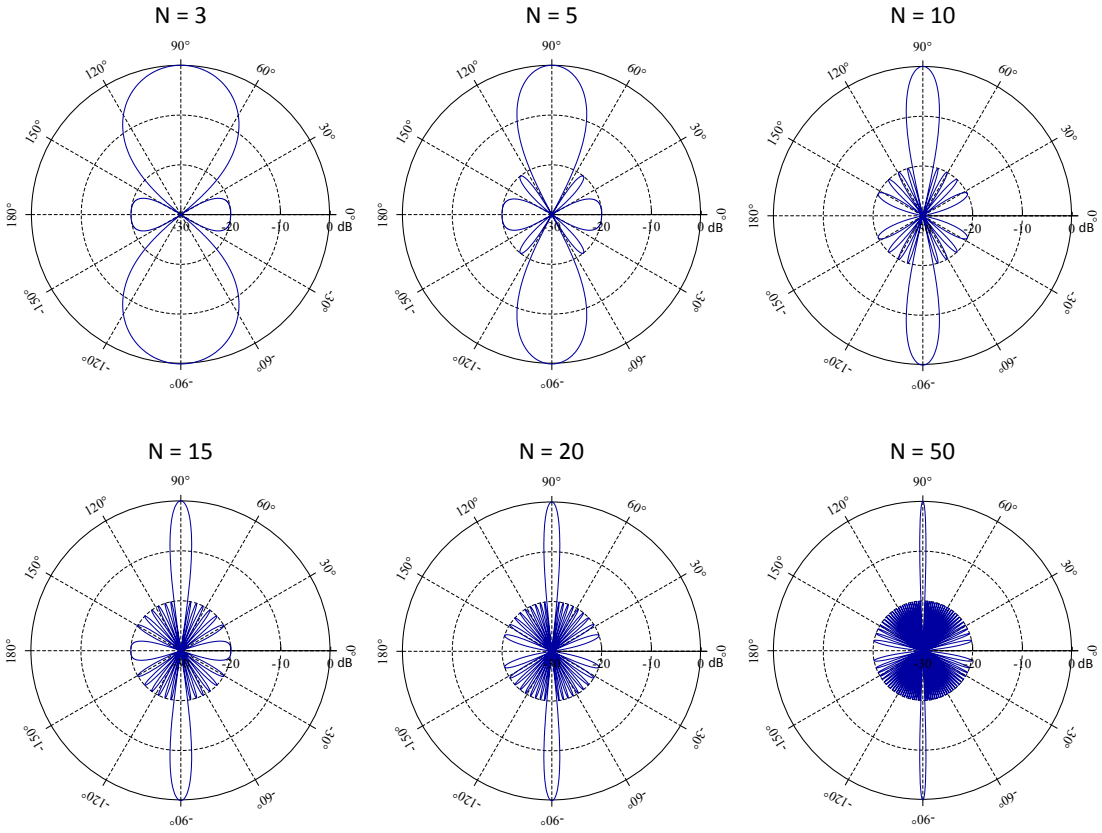
Figure 3.3. AF for different amplitude distributions: uniform, binomial, and Dolph-Tschebyscheff (20 dB SLL). An array with $N = 10$ elements and element spacing $d = \lambda/2$ is used. The amplitude distribution and half-power beamwidth (HPBW) are also given for each distribution.

The Dolph-Tschebyscheff distribution is somewhat of a compromise between the uniform and binomial distribution. When implementing this distribution, the SLL (sidelobe level) is specified, and when the sidelobes are eliminated, it reduces to a binomial distribution. When using this distribution, all sidelobes are at a constant level as opposed to the uniform distribution where the SLL decreases for lobes farther from the main beam. Specifying lower sidelobes widens the beamwidth and increases the disparity between individual element amplitudes. Therefore, the Dolph-Tschebyscheff distribution allows a compromise to be chosen between SLL, beamwidth, and amplitude disparity.

Number of Elements

The number of elements greatly affects a phased array's performance and cost. Adding elements to an array reduces its beamwidth, and an array with a large number of elements can produce a very narrow beamwidth. This is illustrated in Fig. 3.4.

However, more elements means that the system will be larger and more expensive. Phase-shifters can be quite expensive, and since each element must have its own phase-shifter and controlling mechanism, the expense can add up very quickly.



Dolph-Tschebyscheff d = λ/2

N	3	5	10	15	20	50
HPBW	**40.4°**	**23.6°**	**11.2°**	**7.2°**	**5.4°**	**2.1°**

Figure 3.4. Illustration of the effect of N (number of elements) on the AF's beamwidth. A Dolph-Tschebyscheff (20 dB SLL) array with element spacing $d = \lambda/2$ is used for comparison.

Element Spacing

As previously mentioned, close element spacing will delay the onset of grating lobes as illustrated in Fig. 3.5; and (3.17) can be used to find the maximum element spacing (for a given scan angle) which will prevent grating lobes.

The tradeoff is that closer element spacing will also increase the beamwidth as illustrated in Fig. 3.6. Therefore close element spacing is desired to eliminate grating lobes, but large element spacing is desirable for a narrow beamwidth. Also, the element spacing is expressed as a function of wavelength, so if the array is to operate at multiple frequencies, the effective element separation will be different for each frequency. This effect should be taken into consideration in the design.

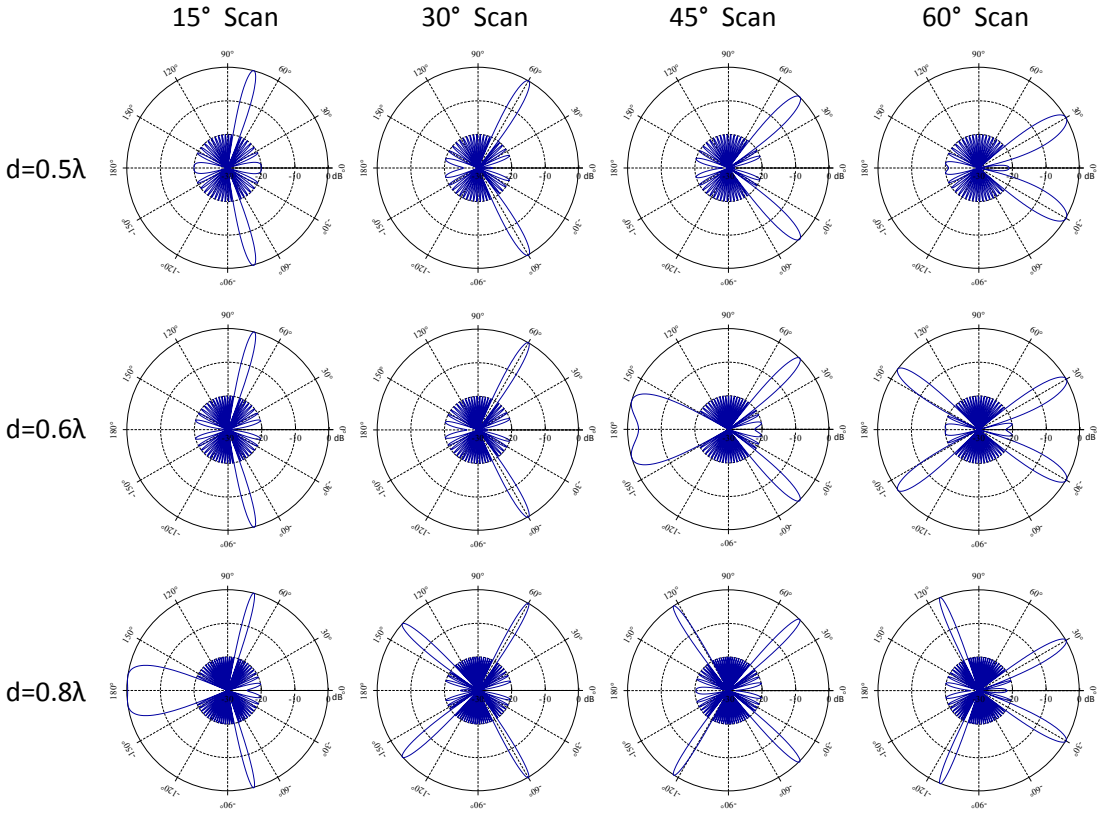


Figure 3.5. Illustration of the effect of d (element spacing) on grating lobes in the AF. A Dolph-Tschebyscheff (20 dB SLL) array with $N = 20$ elements is used for comparison.

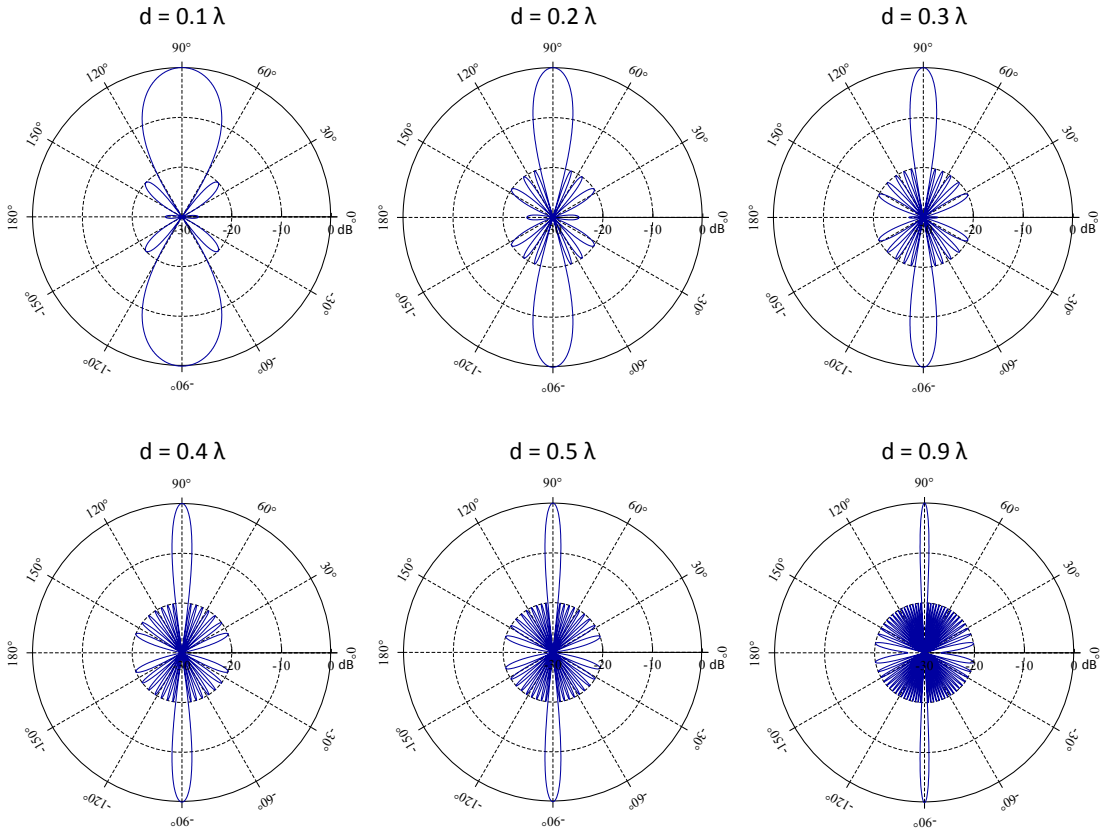


Figure 3.6. Illustration of the effect of *d* (element spacing) on AF beamwidth. A Dolph-Tschebyscheff (20 dB SLL) array with $N = 20$ elements is used for comparison.

Phase-Shifter Bits

The effect that the number of phase-shifter bits has on phase quantization grating lobes has been previously discussed and quantified in (3.28)-(3.29) and graphed in Fig. 3.2. A phase-shifter with many bits reduces the size of the quantization lobes, but it is in general more expensive than one with fewer bits. Increasing the number of bits also increases control complexity and cost.

Scan Range Considerations

The AF's gain is constant across all scan angles. However, since practical antennas are not isotropic, the single-element antenna's gain will decrease as the observation angle diverges from broadside. Since the array pattern is determined by the product of the array factor and the single-element pattern, the array's gain will decrease with increasing scan angles. If AF beamwidth is wide, the AF gain at angles which are lower than the intended scan angle can be large enough that when multiplied by the larger single-element gain, the array's gain will actually be larger at smaller angles than it is at the intended scan angle. This will effectively create a smaller scan angle than intended and can decrease the array's maximum scan angle.

Fig. 3.7 illustrates this by showing different scan angles for arrays of various sizes. The AF and single-element pattern are illustrated in thin red and blue, respectively. The product of these gives the array pattern which is shown in thick black. For a 4-element array, a 30° scan works well, but already the array points to an angle slightly less than 30°. For a 45° scan, the array points to an angle slightly larger than

30°, and for a 75° scan, the sidelobes have exceeded the magnitude of the main beam. The 8-element array scans significantly better, and the 16-element array only starts to lose angle at the 75° scan angle.

It has been shown that the number of elements in an array has a significant impact on the array's maximum scan angle and performance. Hence, the required scan angle and beamwidth should be taken into account when designing a phased array.

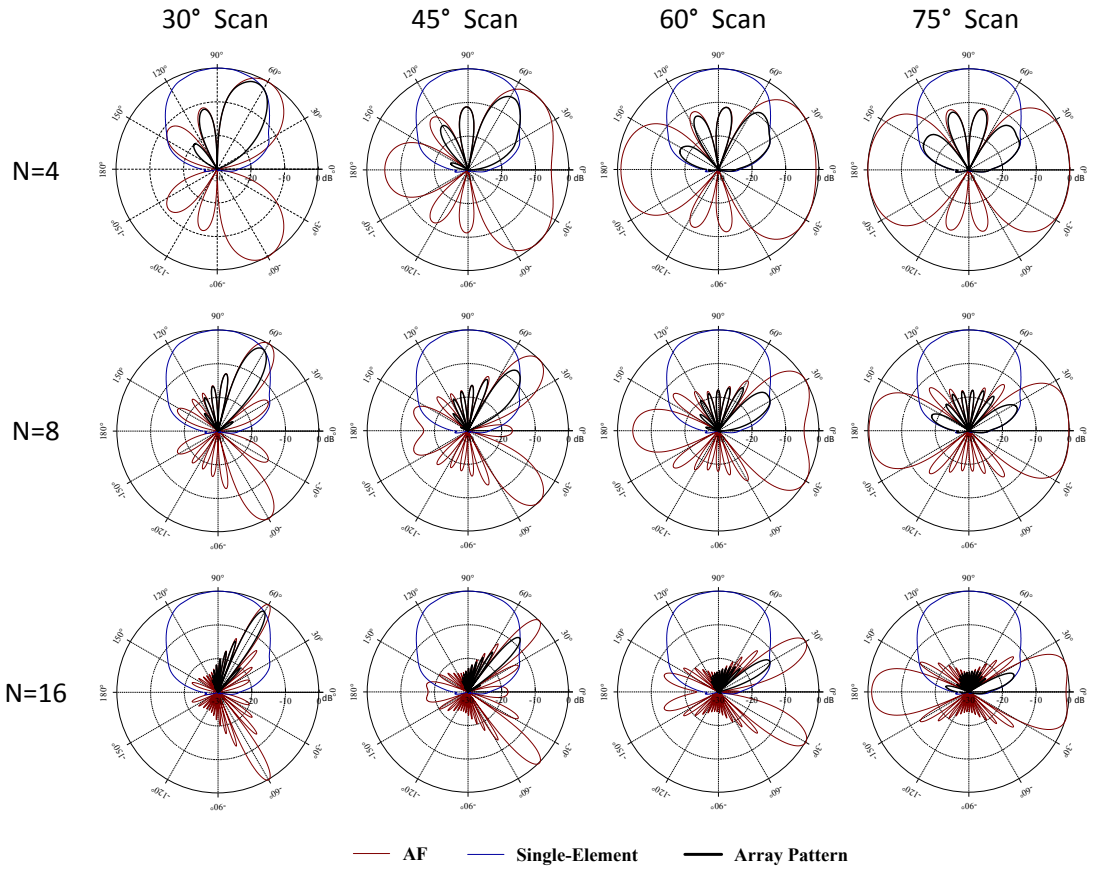


Figure 3.7. Illustration of the effect of N (number of elements) on scan angle. The AF and Single-Element patterns are shown in red and blue, respectively. The Array Pattern is the product of these two and is shown in black. The measured single-element E-plane pattern from the previous section (for Feed 1 at 10 GHz) and a uniform array with $d = \lambda/2$ are used to generate the plots.

4. PHASED ARRAY DEVELOPMENT

A. Antenna Array

The previously designed single-element antenna is to be used in a 4x1 array configuration. A uniform amplitude arrangement is chosen at this stage to simplify the design and avoid the need for amplifiers or attenuators. At a later date, computer-controlled amplifiers can be incorporated in the design to allow various amplitude schemes to be easily realized. The element spacing is chosen as 0.5λ at 10 GHz (1.5 cm) to minimize the effect of grating lobes while still maintaining a reasonably narrow beamwidth. This equates to 0.425λ at 8.5 GHz and 0.575λ at 11.5 GHz. A diagram of the array configuration can be seen in Fig. 4.1.

The array is fabricated in a manner similar to the method previously described for the single-element, with care being taken to ensure proper alignment; the fabricated array can be seen in Fig. 4.2. The four feeds on one side of the array will be combined into port 1, and the four feeds on the other side will be combined into port 2. The port numbering is defined from the numbering paradigm set by the single-element feed numbering (i.e., port 1 is created from the top feeds and port 2 is created from the bottom feeds).

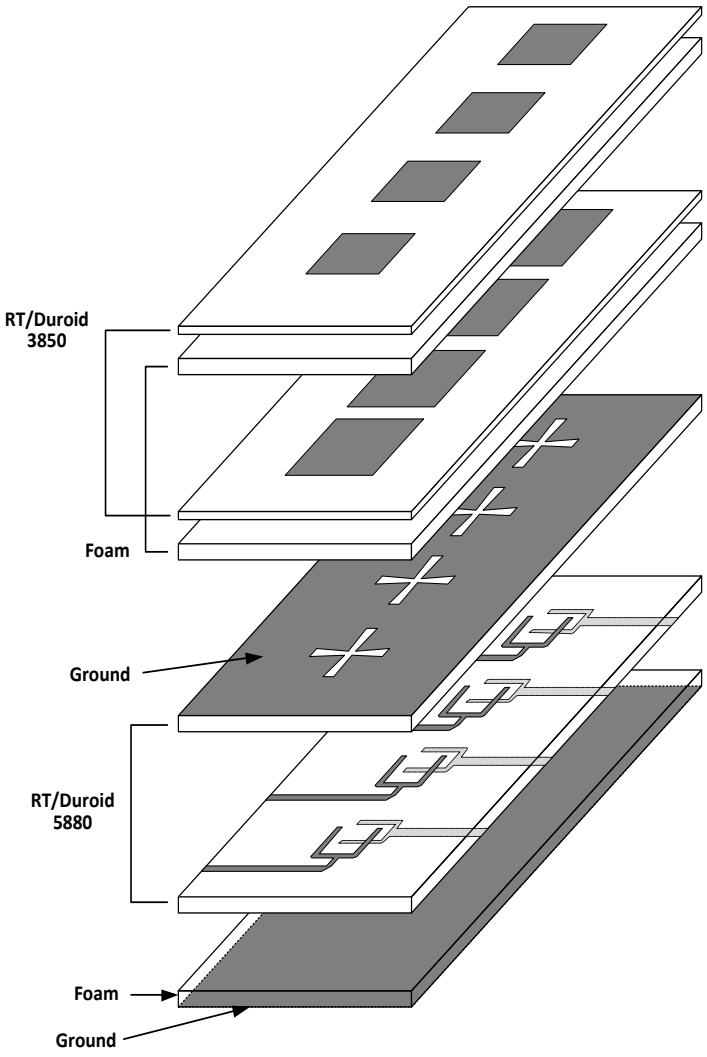


Figure 4.1. 4-element array design (stacked-patch, aperture-fed, dual-polarized stripline).

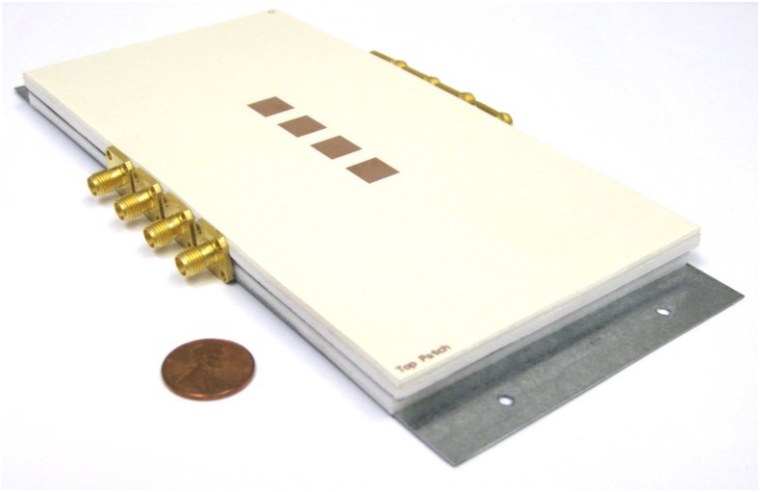
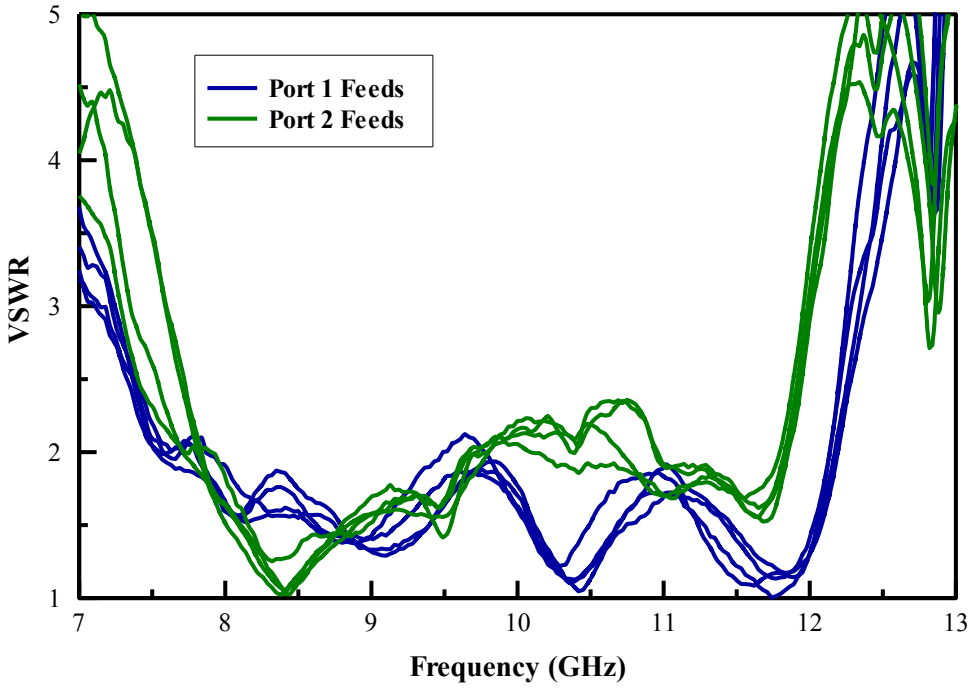
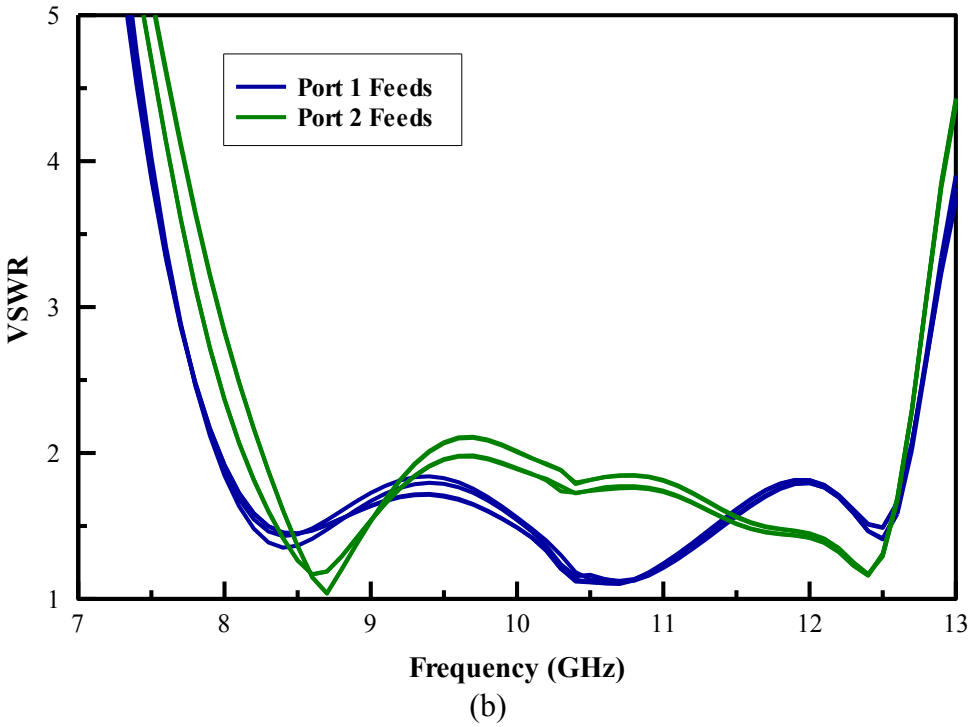


Figure 4.2. Photograph of the antenna array.

An Agilent 8510C network analyzer was used to measure the VSWR of the feeds as well as the isolation between the feeds. These measured values as well as simulated values can be seen in Figs. 4.3-4.4. Ignoring some small overruns, port 1's feeds have a simulated VSWR bandwidth of 45% (8 – 12.7 GHz) (f_0 = 10.4 GHz) and a measured VSWR bandwidth of 43% (7.9 – 12.2 GHz) (f_0 = 10.1 GHz). Port 2's feeds have a simulated VSWR bandwidth of 42% (8.3 – 12.7 GHz) (f_0 = 10.5 GHz) and a measured VSWR bandwidth of 41% (7.8 – 11.8 GHz) (f_0 = 9.8 GHz). This shift in bandwidth between simulated and measured values is similar to what was seen with the single-element and is the reason for the redesign. This places the center frequency close to 10 GHz. The simulated isolation between the ports' feeds is greater than 20 dB, and the measured isolation is greater than 18 dB throughout the VSWR bandwidth.

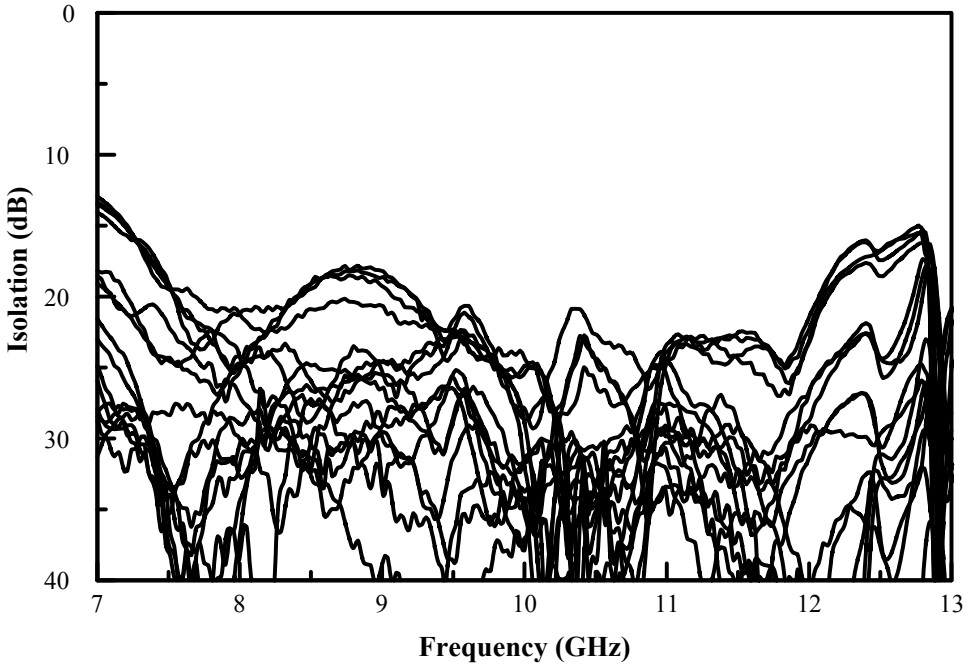


(a)

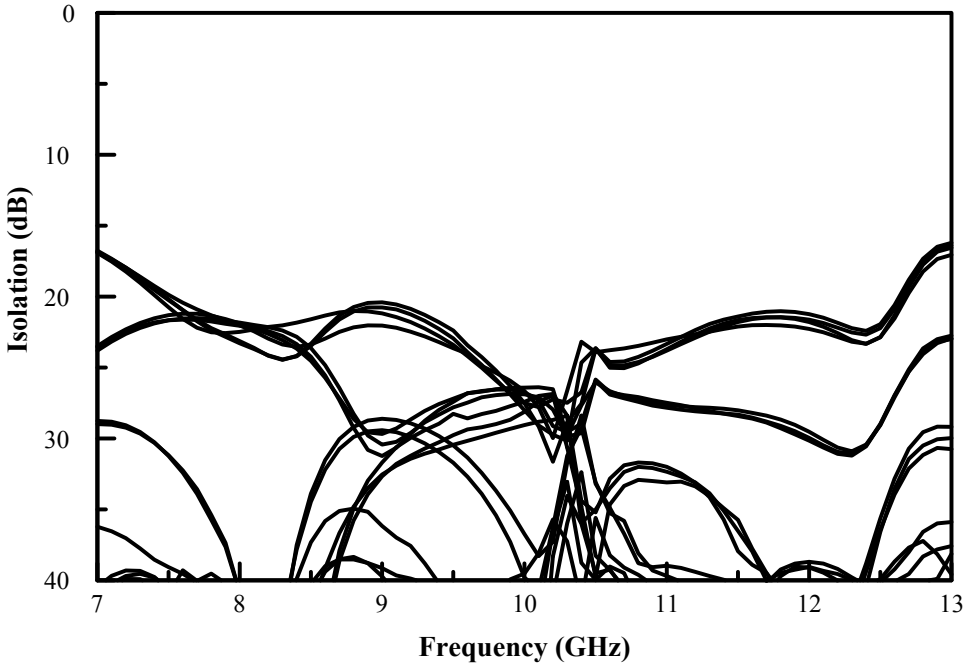


(b)

Figure 4.3. VSWR of the array feeds. (a) Measured and (b) simulated.



(a)



(b)

Figure 4.4. Isolation of the array feeds. (a) Measured and (b) simulated.

B. Power Divider

A 1-to-4 power divider must be used to evenly divide power between the four feeds of each port. Ideally, each output of the power divider will have the same amplitude and phase shift. However, the phase-shifters can be used to adjust for any difference in phase, but since variable amplifiers (or attenuators) are not used in this design, there is no way to adjust for an amplitude variation.

A corporate feed [24] arrangement is used to divide power since this can support a wide bandwidth. In this arrangement, a mitered T-junction [25] will be used to split the power into two, and then mitered T-junctions will be used to split that power again.

In addition to splitting power, the mitered T-junction will change the impedance from Z_0 at the input to $2*Z_0$ at the two outputs. This change in impedance must be accounted for with a transformer. The transformer will be placed before the T-junction rather than after it so that it changes the impedance from 50 to 25 Ω rather than from 100 to 50 Ω. This will prevent the use of high impedance lines which are narrow and sensitive to fabrication errors. It will also reduce the number of required transformers from six to three.

A single quarter-wave transformer can only support a narrow bandwidth, so it is better to use a multi-section transformer to increase the bandwidth. A two section transformer is sufficient and can be designed from a simple algorithm using a uniform geometric progression as presented in [26]. The characteristic impedance Z_0 is 50 Ω, and the load resistance R_L, which is the required impedance before the T-junction, is 25 Ω. The number of sections N will be 2. The impedance will be transformed into

intermediary impedances before finally reaching the final value. The step between these intermediary impedances is defined as:

$$Step = \left(\frac{R_L}{Z_0}\right)^{1/N} = \left(\frac{25}{50}\right)^{1/2} = 0.7071 \quad . \tag{4.1}$$

Then the intermediary impedances are defined as:

$$Z_i(1) = Step \text{ x } Z_0 = 35.6\ \Omega \tag{4.2}$$

$$Z_i(2) = Step \text{ x } Z_i(1) = 25\ \Omega \quad . \tag{4.3}$$

Finally, the characteristic impedance of each quarter-wave section can be defined as:

$$Z(1) = \sqrt{Z_i(0)Z_i(1)} = 42.04\ \Omega \tag{4.4}$$

$$Z(2) = \sqrt{Z_i(1)Z_i(2)} = 29.73\ \Omega \quad . \tag{4.5}$$

The power divider design can be seen in Fig. 4.5, and a photograph can be seen in Fig. 4.6. Note that there is not a 25 Ω line in the design because the input to the mitered T-junction is placed immediately following the conversion to 25 Ω.

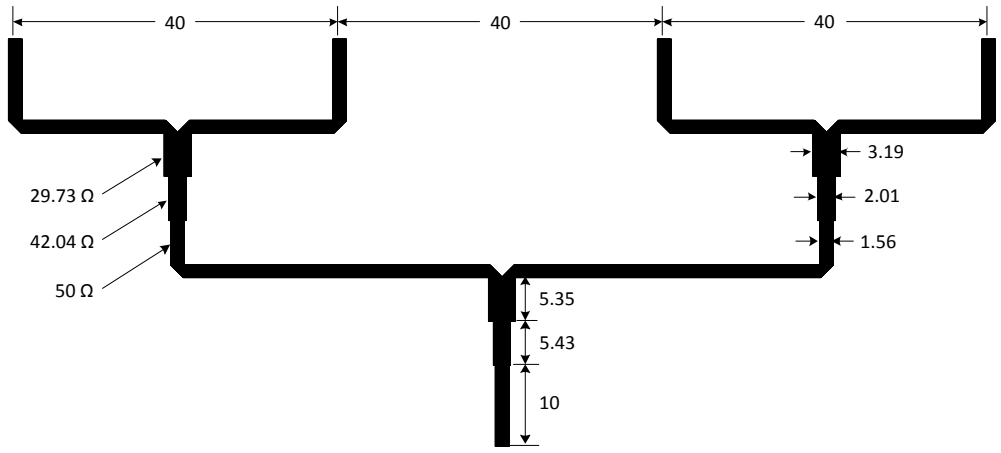


Figure 4.5. Schematic of the power divider (all units in mm).

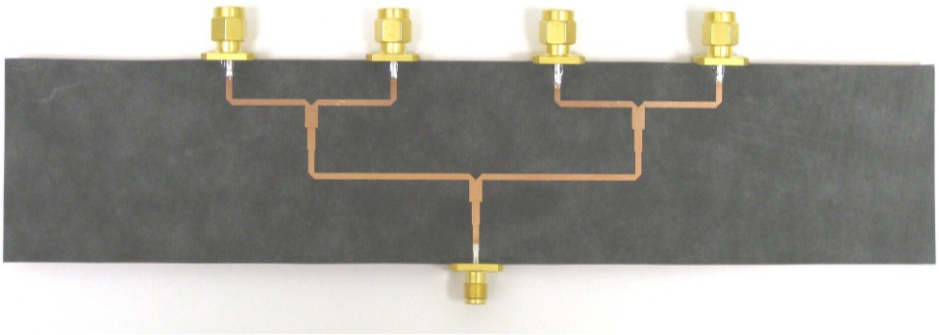
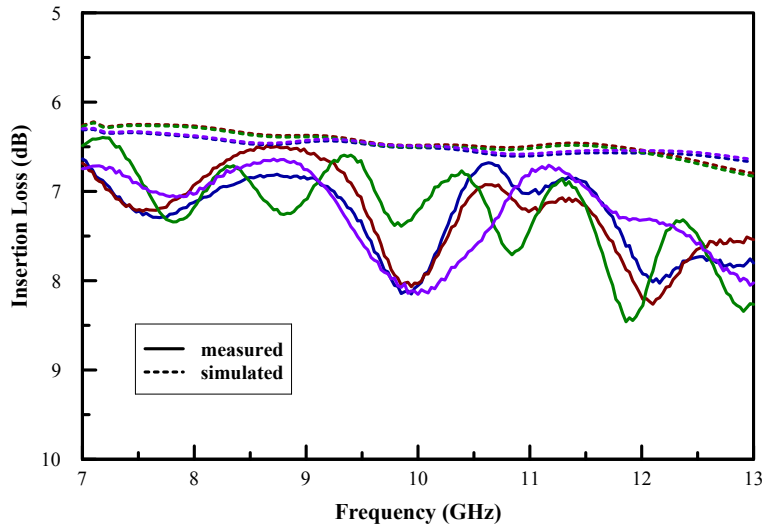


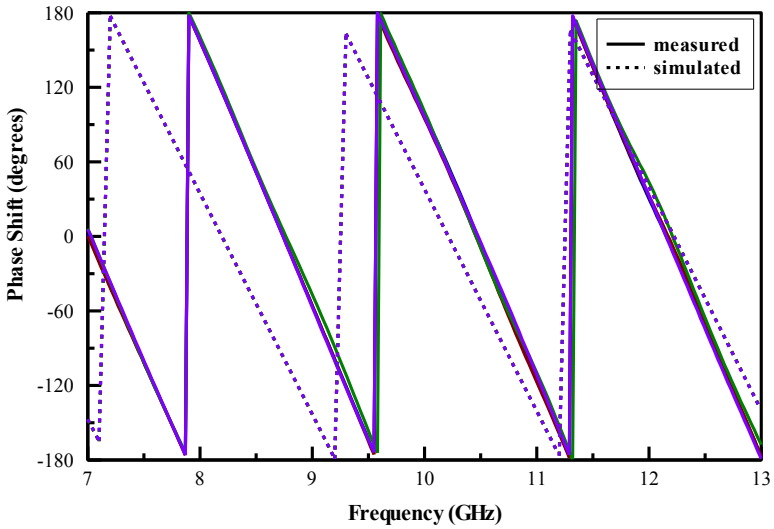
Figure 4.6. Photograph of the power divider.

The insertion loss of the power divider is measured with a network analyzer, and the results are displayed in Fig. 4.7. These are plots of the insertion loss and phase shift from the input to each of the outputs, so ideally the insertion loss and phase shift of all the outputs would be equal to one another. The simulation shows that, at every frequency, the insertion loss of any two outputs differ by less than 0.19 dB, and the phase shifts differ by less than 1°. The measurements show that the insertion losses

differ by less than 1.16 dB, and the phase shifts differ by less than 14°. This error in measured values is most likely due to fabrication errors, namely connector inconsistencies.



(a)



(b)

Figure 4.7. Measured and simulated (a) insertion loss and (b) phase shift of the power divider.

C. Phase-Shifters

Four digital phase-shifters are needed to control the phased array (one shifter for each element), and there are a number of varieties and options available on the market. Some of the phase-shifters are packaged as a sealed unit in a hardened case which is durable but also large and expensive. The other option is a monolithic microwave integrated circuit (MMIC) which is much cheaper and smaller. They are typically a few square millimeters and can be directly integrated into the antenna array. Using MMIC phase-shifters would allow the entire phased array (antenna, phase-shifters, and power divider) to be fabricated together and be very compact. The downside to the MMIC is that they can be very fragile, and if an integrated phase-shifter breaks then the whole array assembly must be discarded and rebuilt. The other downside is that special equipment is required to mount and connect MMICs, and unfortunately this lab does not have access to that equipment.

There is another option for using MMIC phase-shifters which overcomes this lab's inability to mount them. The companies which manufacture MMICs usually also make an evaluation board with the MMIC already mounted on a board and attached to external connections. Using evaluation boards would sacrifice some of the MMICs advantages since it would no longer be possible to fabricate the entire phased array together, and they are more expensive than the standalone MMICs. However, the evaluation boards are still smaller and less expensive than the sealed units. For these reasons, MMIC evaluation boards will be used in the system.

The Hittite HMC543LC4B phase-shifter is chosen for use in this system mainly for its price. It is a 4-bit digital phase-shifter which operates from 8 – 12 GHz. It has 360° coverage, so 4-bits means that the smallest possible adjustment is 22.5°. A photograph of the HMC543LC4B phase-shifter evaluation board can be seen in Fig. 4.8.

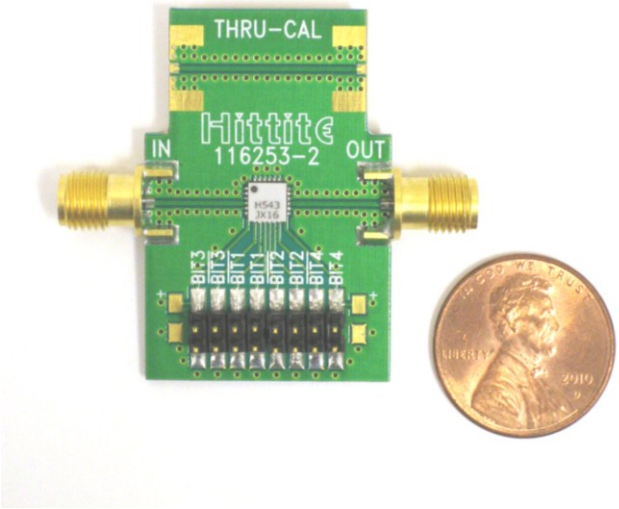


Figure 4.8. Photograph of the Hittite HMC543LC4B phase-shifter evaluation board.

The S-parameters of the phase-shifter are measured with a network analyzer, and the results can be seen in Figs. 4.9-4.10. Each line in the graphs represents a different phase setting. The insertion loss for each phase setting (Fig. 4.9) ranges from 6.3 – 8.3 dB which is a range of 2 dB. However, at any given frequency, the insertion loss varies by less than 1.6 dB, and at 10 GHz it varies by less than 0.6 dB. The variation in insertion loss increases for frequencies farther away from the center frequency of 10 GHz. This insertion loss variation will create a deviation from the

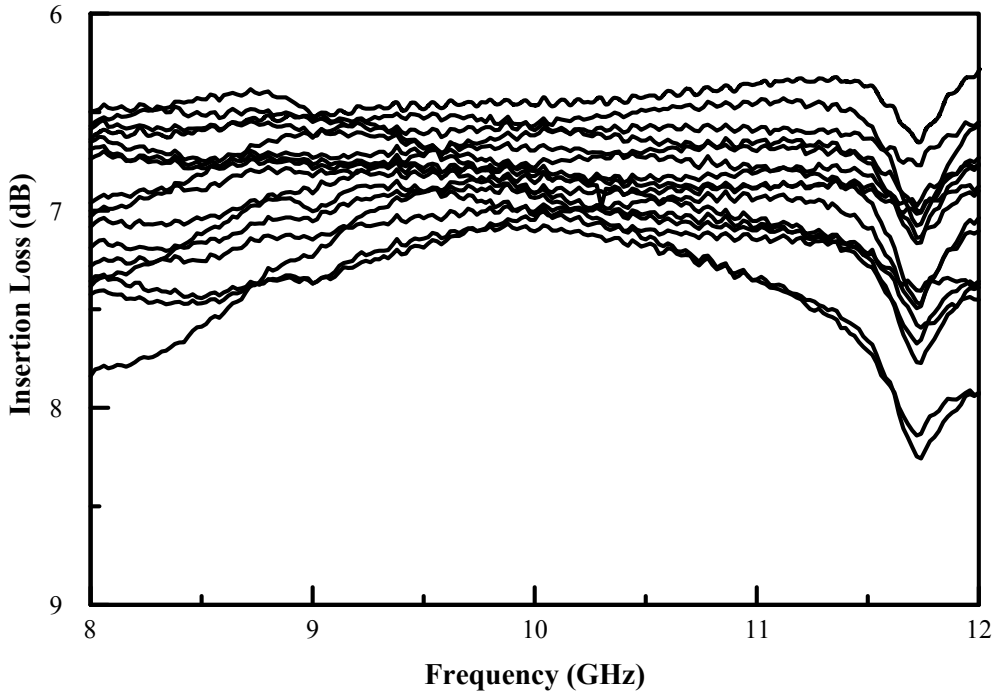


Figure 4.9. Measured insertion loss of the phase-shifter at all phase settings.

planned uniform amplitude distribution, and if variable amplifiers were included in the design, this variation could be accounted for.

Fig. 4.10a illustrates the relative phase shift of each setting where each setting is relative to the 0° state. Ideally each setting would be 22.5° from the previous setting, but some deviation exists. This deviation can be seen in Fig. 4.10b which illustrates the amount that each setting deviates, and it can be seen that there is a maximum deviation of 12°. This phase deviation can be corrected in the calibration. However, the phase shift error of each state is not consistent for all frequencies, and it can vary by as much as 18° (as a function of frequency) for some states. Hence, if the system is calibrated at

one frequency, there will be some phase error introduced at other frequencies by the inconsistencies of the phase-shifter. Of course this can be corrected by recalibrating at other frequencies and using different settings for different frequencies, but if the system is to be used in a broadband application, a single setting should be used.

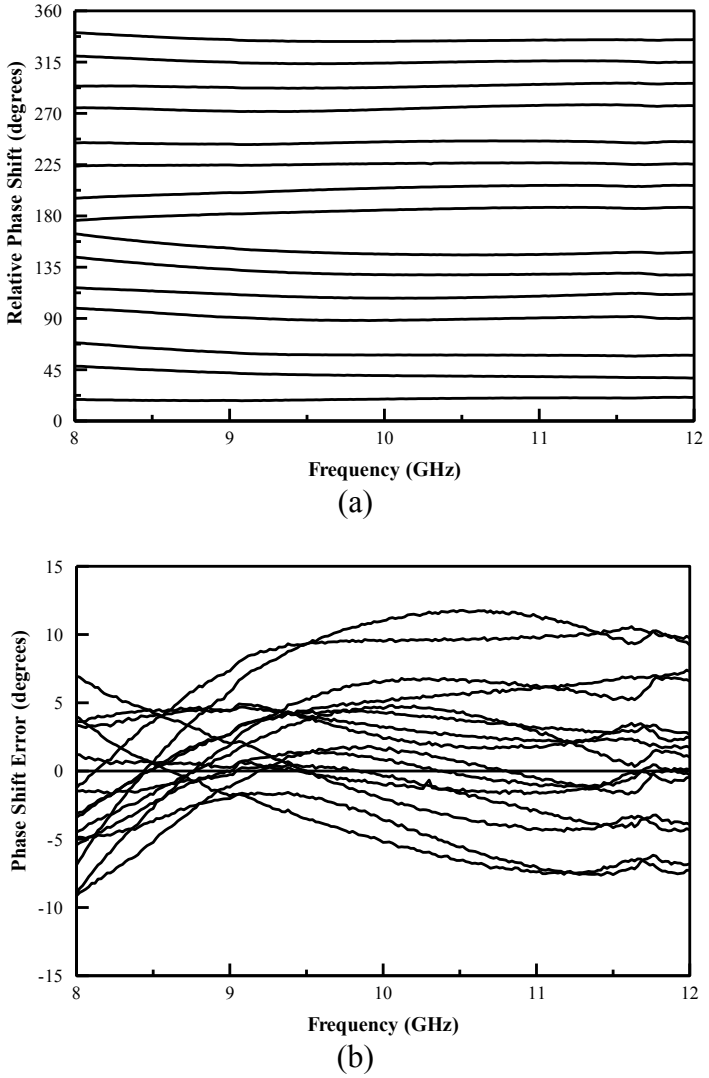


Figure 4.10. Measured (a) relative phase shift of the phase-shifter at all phase settings and (b) phase shift error.

D. Phase-Shifter Control

To steer the phased array, the phase-shifters must be digitally controlled by a computer. There are four 4-bit shifters, so 16 bits of information are required. LabVIEW [27] can be used in conjunction with a Data Acquisition (DAQ) device to provide the 16 bits. An NI USB-6509 Digital I/O device will serve as the DAQ with an NI CB-50 connector box providing the connection terminals. This DAQ provides 96 channels (operating on the 5 V TTL/CMOS standard) which will allow room for the array to grow at a later date. Photographs of the DAQ and connector box can be seen in Fig. 4.11.

(a) (b)

Figure 4.11. Photographs of (a) the NI USB-6509 DAQ and (b) the NI CB-50 connector box.

LabVIEW can be used to convert the desired phase to a 4-bit representation which can be sent to the phase-shifter. The conversion procedure is as follows:

1.) Shift phase to the range 0 - 360° [phase modulo 360]

2.) Convert phase to a 4-bit representation [round(phase/22.5)]

3.) If the result is 16, change this to 0 (16 is an overflow)

Outputs are sent in packets of 8 bits, so two phases should be combined into one 8-bit output as follows:

1.) (4-bit phase 1) + (4-bit phase 2) x 16

where the least significant bits (LSBs) are for shifter 1 and the most significant bits (MSBs) are for shifter 2. This procedure is used to develop an easy to use LabVIEW interface (Fig. 4.12), and the details of the implementation can be seen in the appendix.

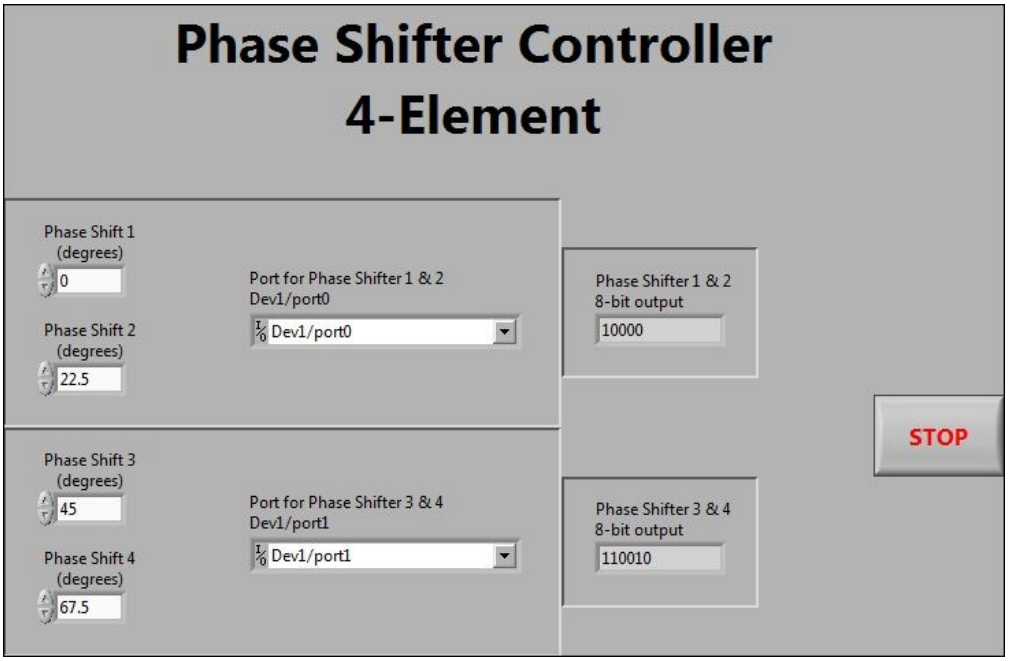


Figure 4.12. LabVIEW interface for controlling four 4-bit phase-shifters.

E. Phase-Shifter Logic Conversion

LabVIEW and the DAQ provide the standard 5 V TTL/CMOS logic, but unfortunately the phase-shifters require a complementary logic of 0/-3V. In this logic scheme, each bit of information requires two control lines (one with 0 V and the other with -3 V). A converting circuit must be designed to convert all 16 control bits into 32 bits of this 0/-3V complementary logic. A circuit design [28] as seen in Fig. 4.13 can be used to convert each bit.

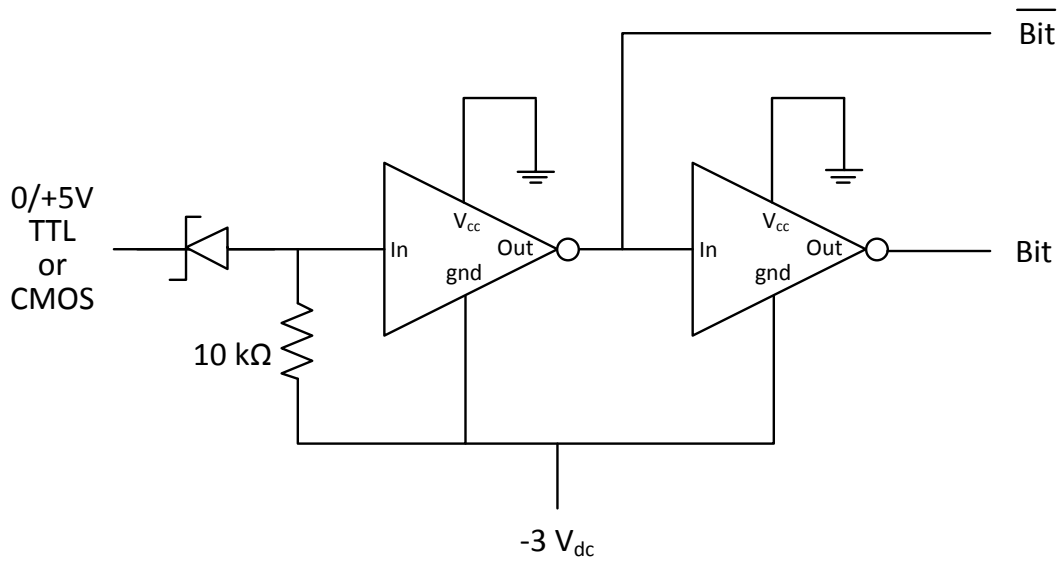
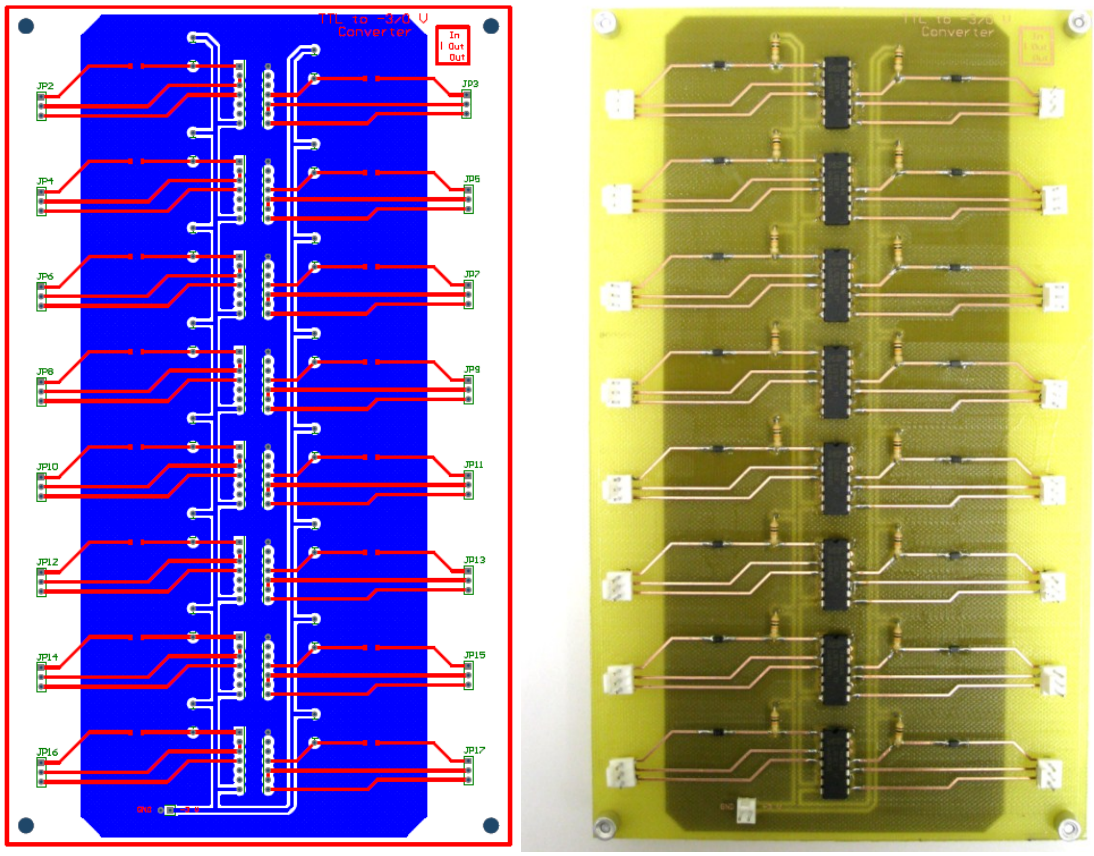


Figure 4.13. Circuit to convert 5 V TTL/CMOS logic to 0/-3V complementary logic.

This circuit contains two inverters (Philips Semiconductors 74HC04), a 10 kΩ resistor, and a zener diode (ON Semiconductor MMSZ4689T1, V_z = 5.1 V, I_{zt} = 50 μA). When the input is 5 V, the diode will conduct, and the 5 V zener voltage drop will cause 0 V to appear at the input of the first inverter. This will cause the inverter to connect its

gnd (which is connected to -3 V) to its output. Then -3 V will appear at the output $\overline{Bit}$, and the second inverter will connect its V_{cc} (which is connected to ground) to its output which is the output *Bit*. This will make $\overline{Bit}$ = -3 V and *Bit* = 0 V. When the input is 0 V, the diode will not conduct, and -3 V will appear at the input to the first inverter. This will cause the inverter to connect its V_{cc} (which is connected to ground) to its output. Then 0 V will appear at the output $\overline{Bit}$, and the second inverter will connect its *gnd* (which is connected to -3 V) to its output which is the output *Bit*. This will make $\overline{Bit}$ = 0 V and *Bit* = -3 V.

This circuit must be implemented 16 times in order to convert all of the bits. The most clean and efficient way to do this is with a Printed Circuit Board (PCB) which can be designed using Protel DXP 2004 [29]. The input and outputs for each bit are grouped together and placed on the left and right sides of the PCB. A connection for ground and a -3 V_{dc} source are placed at the bottom, and the inverter DIPs run down the center. The grounds and the -3 V_{dc} leads are connected on the backside of the board. The DXP schematic and a photograph of the assembled PCB can be seen in Fig. 4.14.



(a) (b)

Figure 4.14. PCB to convert 5 V TTL/CMOS logic to 0/-3V complementary logic. (a) Schematic and (b) photograph.

F. Phased Array Assembly

The antenna array, phase-shifters, and power divider are connected and fixed together using a bracket as illustrated in Fig. 4.15. Each polarization will be measured separately with the feeds for the other polarization terminated with 50 Ω loads as shown in the photograph. Three-inch coaxial cables (Mini-Circuits Hand-Flex 086-3SM+) are used to connect the antenna array to the phase-shifters since they can easily be formed to the necessary shape, and the phase-shifters and power divider are directly connected with SMA connectors.

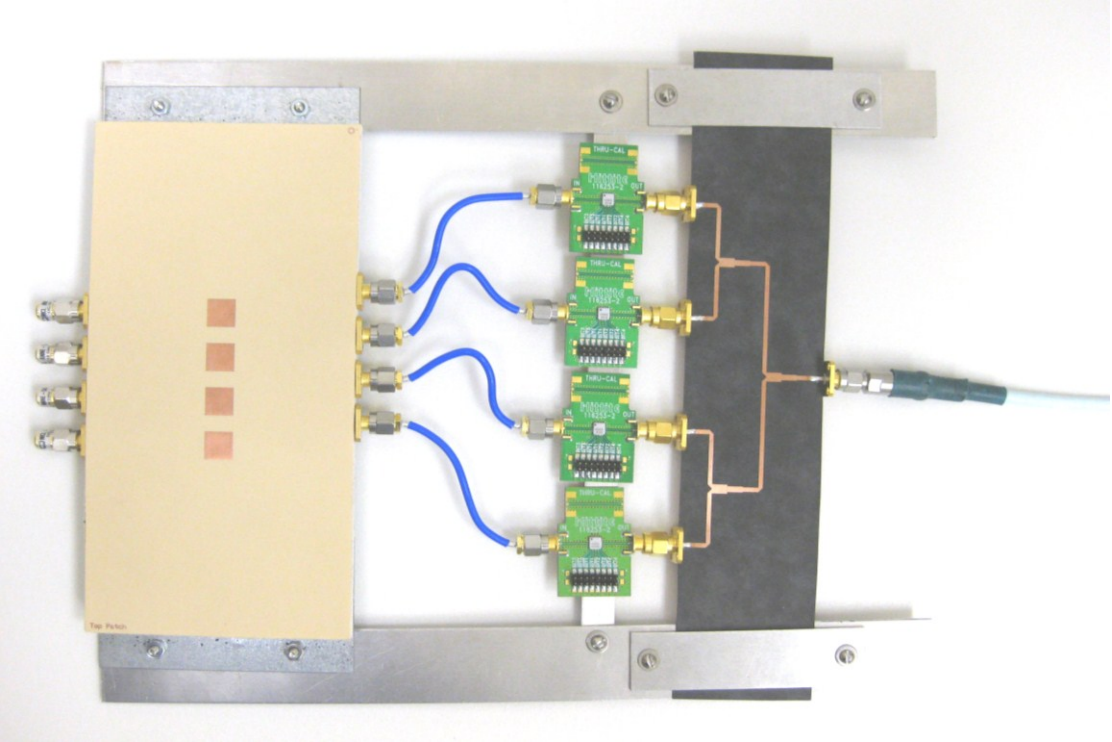


Figure 4.15. Photograph of the array assembly (antenna array, phase-shifters, and power divider) fixed together by a bracket.

The phase-shifters must be wired to the logic converter circuit and the LabVIEW DAQ. Due to the fragility of the phase-shifters, a protective shield is built to protect them from harmful external disturbances as shown in Fig. 4.16. The array assembly is surrounded by absorber to prevent reflections; this is done by placing the assembly in a large hollowed out chunk of absorber and then covering it with layers of absorber as shown in Fig. 4.17.

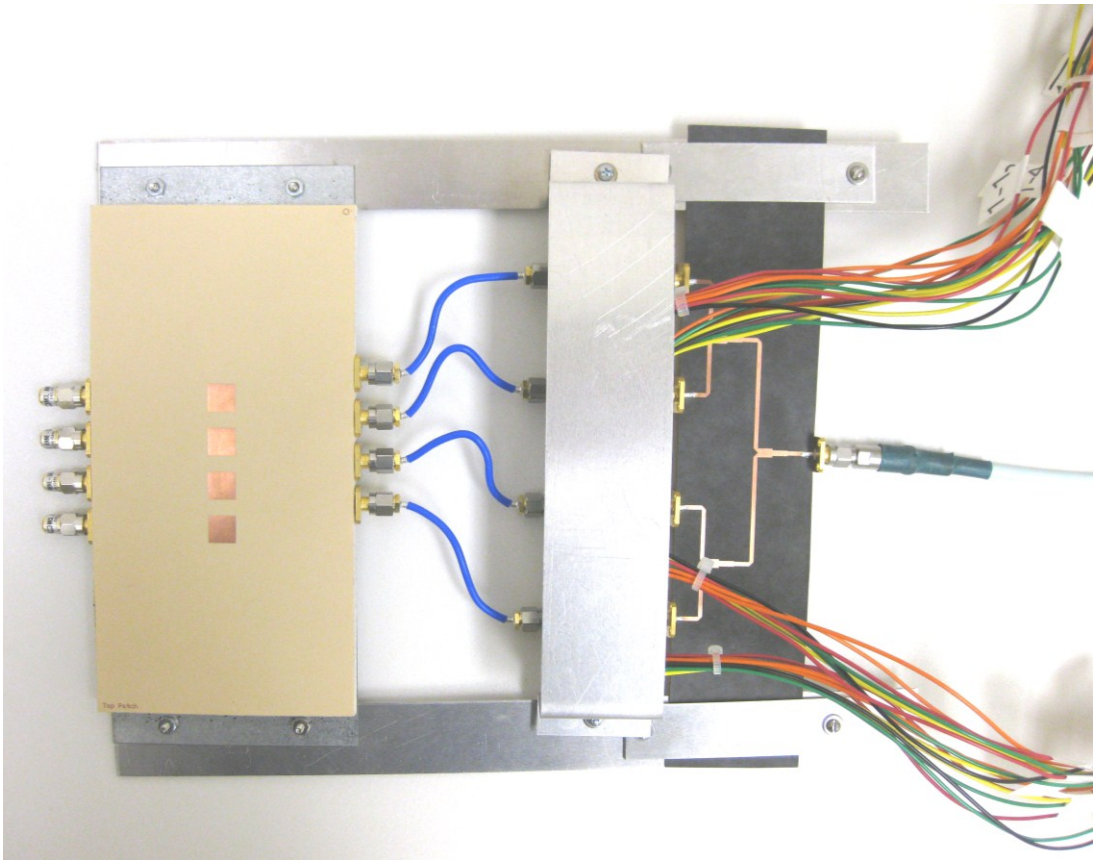
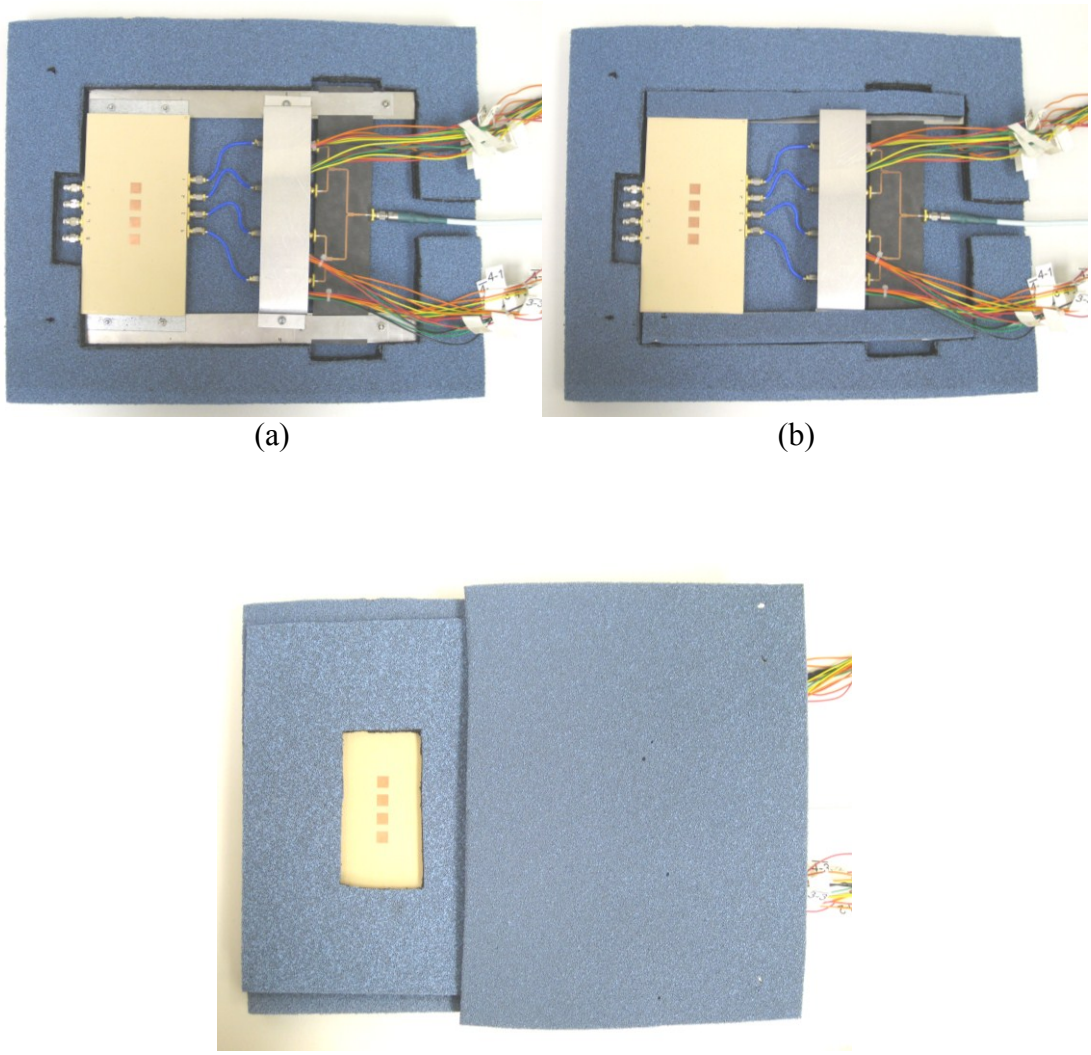


Figure 4.16. Photograph of the array assembly with a protective shield over the phase-shifters.



(a) (b)

(c)

Figure 4.17. Photograph of the array assembly. (a) Placed in a hollowed out chunk of absorber, (b) protected by pieces of absorber, and (c) covered with layers of absorber.

To operate the phased array and take measurements in the anechoic chamber, the array assembly (Fig. 4.17) is mounted inside the chamber, and the wires controlling the phase-shifters are run under the chamber door so that the phase-shifter controlling equipment does not disturb the measurements. These wires are connected to the logic converter circuit which is connected to the connector box which is connected to the DAQ which is connected to a computer running LabVIEW. This arrangement is illustrated in Fig. 4.18, and a photograph can be seen in Fig. 4.19.

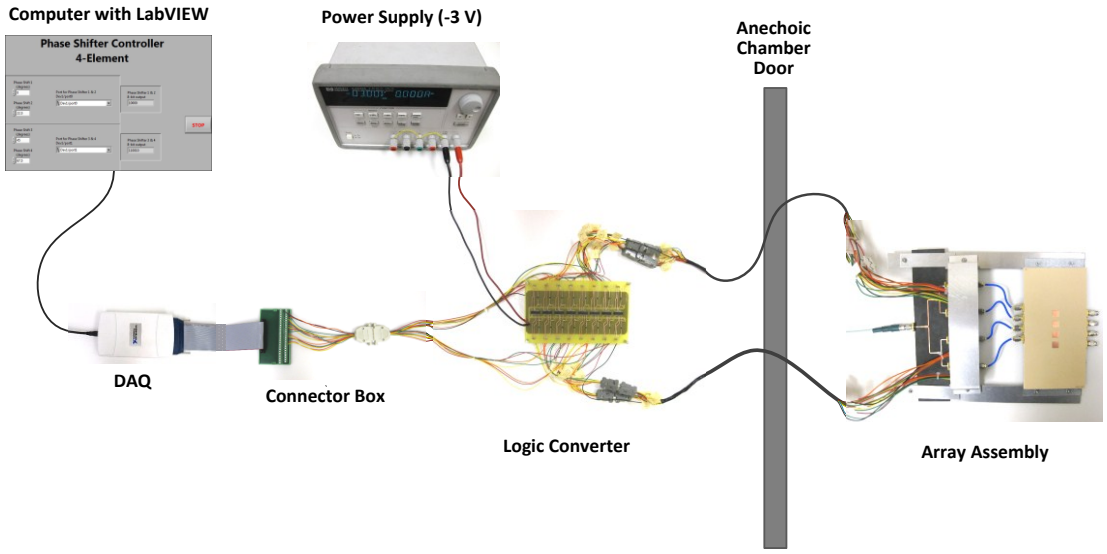


Figure 4.18. Diagram of the phased array system setup.

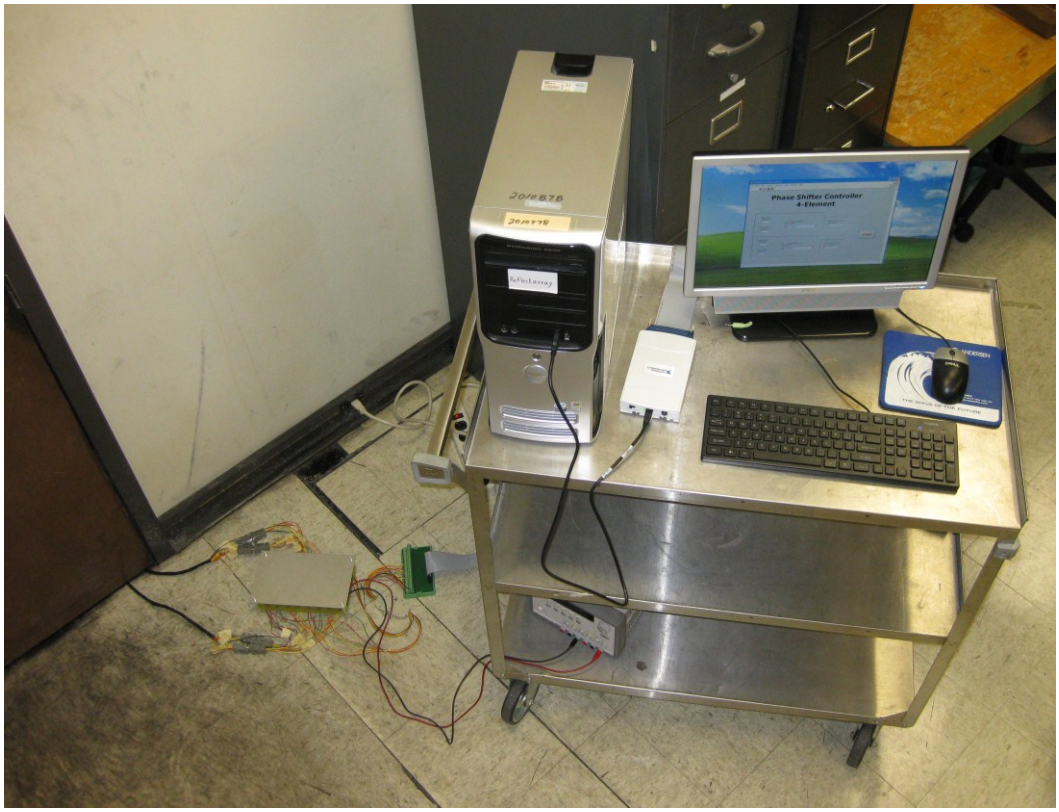


Figure 4.19. Photograph of the phased array system setup.

G. Phased Array Calibration

Before the phased array can be used for the first time, it must be calibrated. Unequal line lengths, phase-shifter errors, and mutual coupling [30] prevent simply offsetting each phase-shifter by identical phases to create the desired scan angle.

Unequal line lengths have the most dramatic effect on calibration. A simple method to calibrate the array is to merely measure the amount of phase shift that the input signal undergoes while traveling through the power divider, phase-shifter, and coaxial line on its way to each feed. The difference in measured phase shift for each

feed can then be subtracted from the desired shift of that feed. The measured phase shifts for lines 1 – 4 are -100°, 65°, 55°, and 60° respectively. There is a difference in phase shift between lines 1 and 4 of 165° which is quite large. However, using this calibration method, the effect of these unequal line lengths can be tuned out. One drawback of this technique is that it does not take into account the phase-shifter errors of each state (as seen in Fig. 4.10b) or mutual coupling [30].

All effects can be accounted for if a far-field phase-alignment calibration approach is used. For this calibration technique, the phased array system is assembled in an anechoic chamber (as seen in Fig. 4.18), and the antenna is oriented at broadside for the duration of the calibration. Each antenna element must be isolated by terminating the other elements' feeds. A far-field gain measurement is then made for all phase states of that element. This is repeated by then isolating the other elements, one at a time, and repeating the measurements for each element. Once a table containing the phase shift of each element for all phase states has been made, the phase settings to be used can be obtained from the calibration table. For example, if a scan angle of 0° is required, phase settings for each shifter should be chosen such that their values in the calibration table are equal. If the array is to be scanned to an angle θ_0, then the phase β between array elements should be

$$\beta = -kd \sin \theta_0 \tag{4.6}$$

where d is the separation between elements and k is the wave number. Then phase settings for each shifter should be chosen such that their values in the calibration table are x, x+β, x+2β, x+3β where x is any number. A plot of the data from one of the calibrations can be seen in Fig. 4.20.

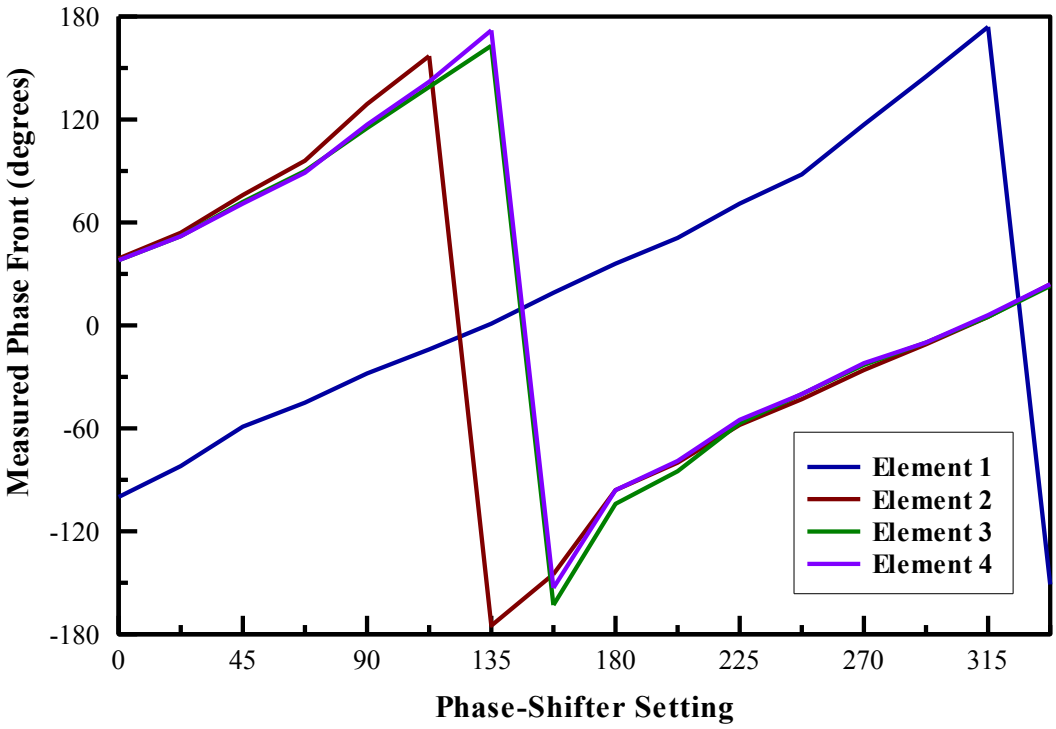


Figure 4.20. Plot of a typical calibration table for the four elements.

There is one small problem with using the far-field phase-alignment method for this phased array. The elements not under test must be terminated, and since variable amplifiers are not used, the only way to do this without disturbing the antenna position is to remove the phase-shifter control lines. This effectively isolates the unwanted elements, but it does this by reflecting power from the phase-shifter rather than

terminating the power. The reflected power from the unwanted elements will combine and interfere with the transmission to the element-under-test. The goal is to isolate the element-under-test and measure it under the same conditions that will be encountered in normal operation, but these reflections disturb that normal operating condition. However, the phase difference from terminating the unwanted elements with a 50 Ω load and unplugging their phase-shifters was measured to be less than 9°, and the variation of this change between elements was measured to be less than 4°. Therefore, the error introduced by unplugging the unwanted elements, rather than terminating them, is minimal.

A comparison of the far-field phase-alignment calibration method and the line length adjustment calibration method is illustrated for a 0° scan angle in Fig. 4.21. Both calibration methods work, but the far-field phase-alignment method gives a higher gain and a nicer beam shape. Consequently, the far-field phase-alignment method will be used.

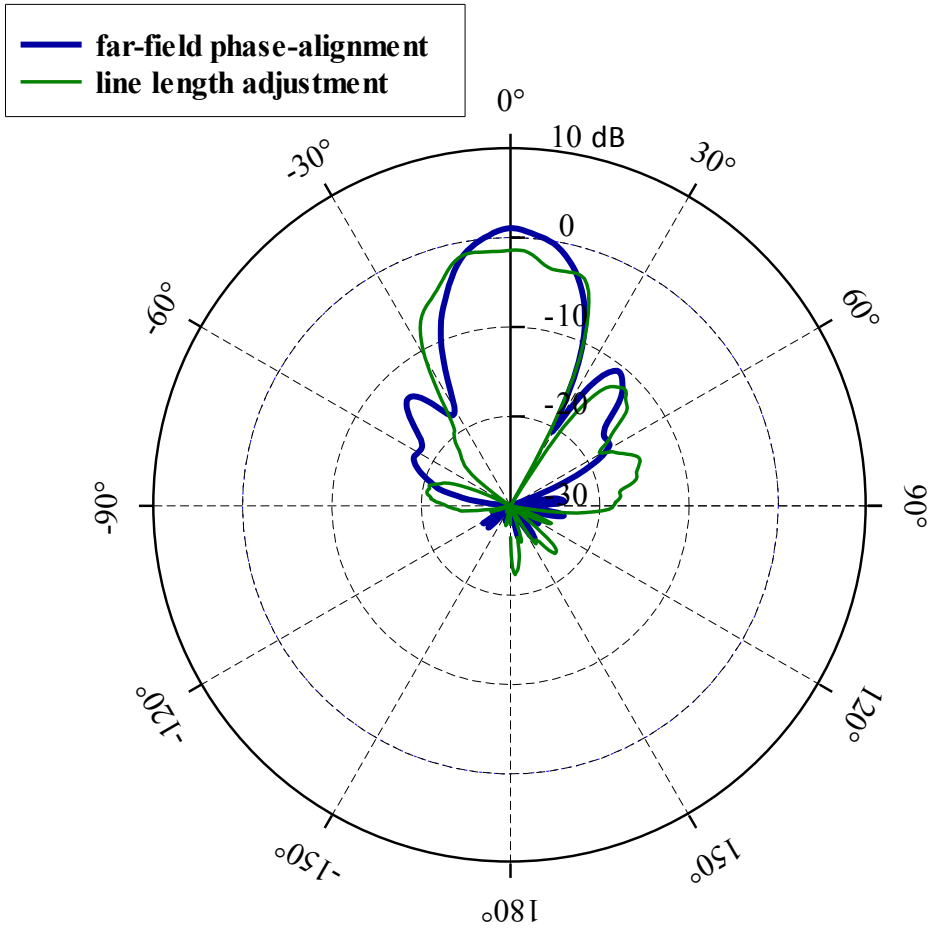
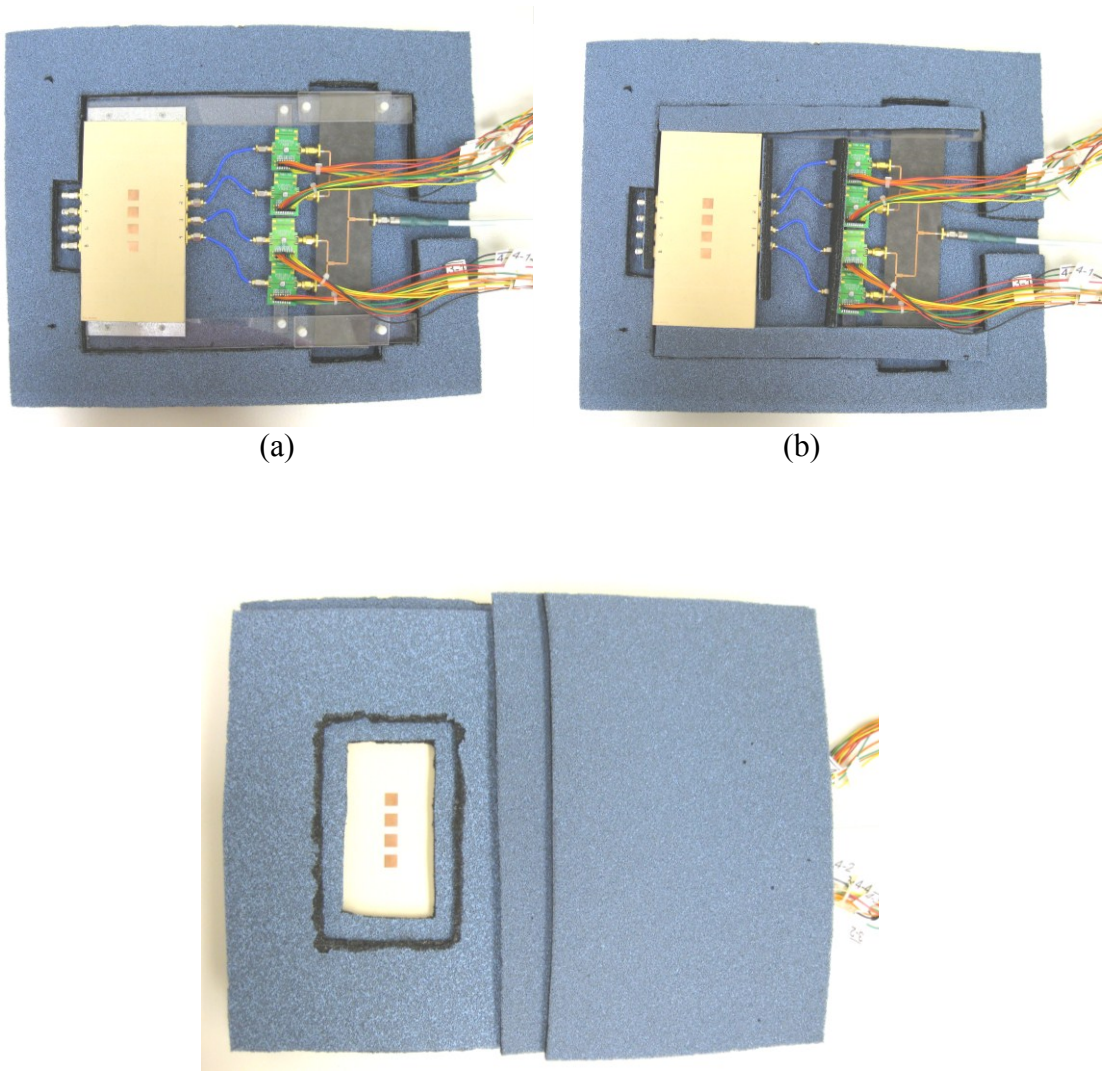


Figure 4.21. Comparison of two phased array calibration methods (the far-field phase-alignment method and the line length adjustment method). Measurement is for a 0° scan angle.

H. Pattern Measurements

The phased array system (Fig. 4.18) is calibrated at 10 GHz, and the pattern is measured from 8.5 to 12 GHz in increments of 0.5 GHz and for scan angles from 0 to 90° in increments of 10°. The array scans as intended and works fairly well, but there are a few abnormalities with the pattern and scan loss. The gain oscillates some with observation angle, and peak gain can sometimes increase with increasing scan angle. This could be caused by a number of factors such as: resonant modes caused by reflection and interaction with array assembly metal, phase quantization errors, calibration errors, and phase-shifter insertion loss variation with phase setting.

The effect from resonant modes caused by reflection and interaction with array assembly metal can be reduced by replacing the array assembly mounting structure with lexan and nylon screws as illustrated in Fig. 4.22a. Also, adding more absorber around the SMA connectors (Fig. 4.22b) and another layer of absorber on top (Fig. 4.22c) can help to prevent and attenuate any reflected waves.



(a) (b)

(c)

Figure 4.22. Photograph of the array assembly with lexan and nylon screws. (a) Placed in a hollowed out chunk of absorber, (b) protected by pieces of absorber, and (c) covered with layers of absorber.

The phase quantization error can be reduced by strategically choosing the reference element which will result in minimum phase quantization error rather than choosing this reference element at random. Since the phase-shifter states are not separated by exactly 22.5°, reference element choice can make a significant difference in phase quantization error. The optimum phase settings for a given phase β can be found using a Matlab [31] program which can be seen in the appendix. If a reference element is chosen at random, the phase error from each element is usually less than 5-12°. However, if the Matlab program is used, the phase error from each element is usually less than 1-3°. This will increase the calibration accuracy and should more correctly align the phase fronts of the elements.

The revamped phased array assembly seen in Fig. 4.22 is calibrated at 10 GHz, and phase-shifter settings are chosen using the Matlab program. The array pattern is then measured as before. The results are similar, but the gain oscillation is reduced, and the scan loss is closer to expectation. Some of the measurements can be seen in Figs. 4.23-4.28.

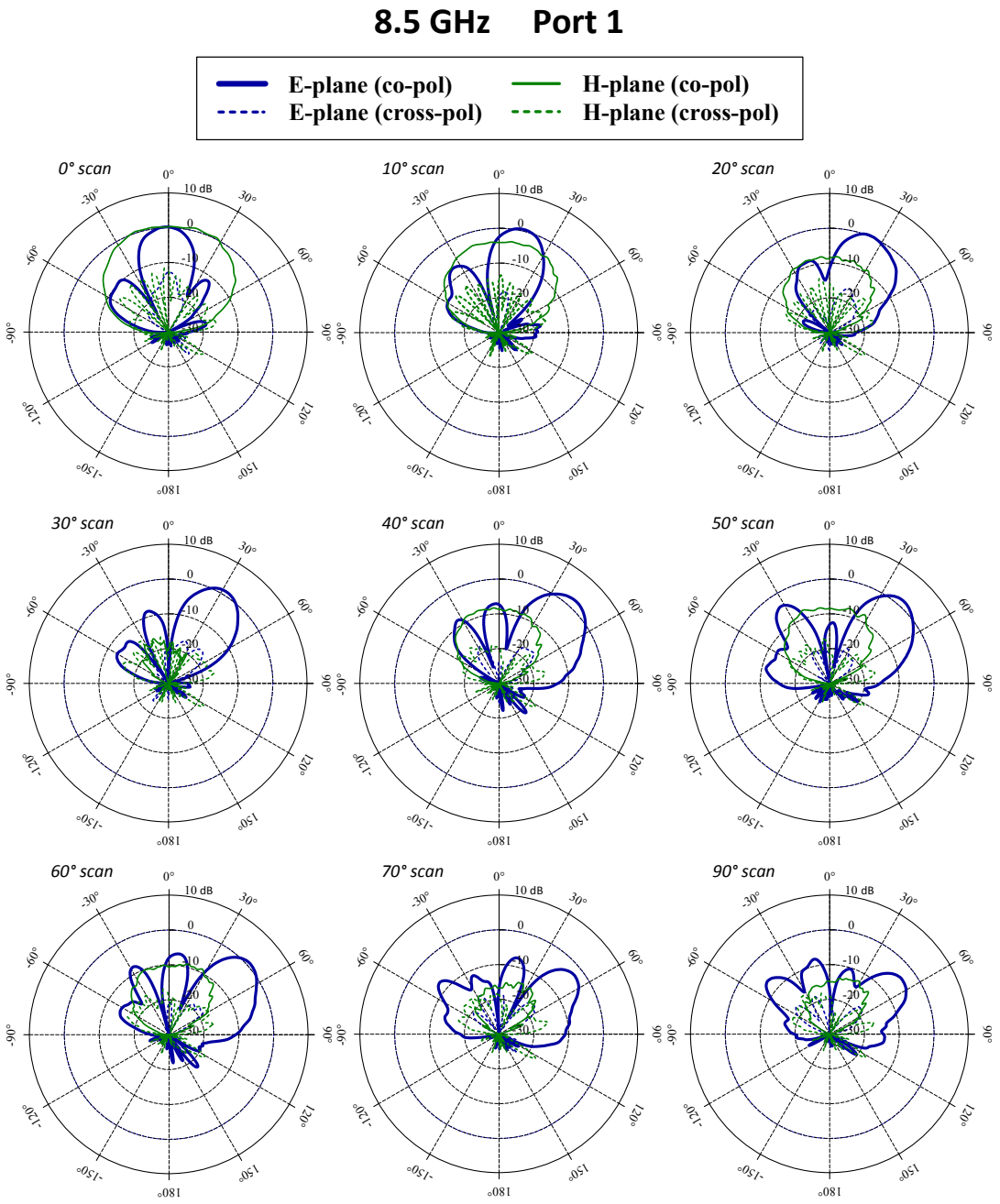


Figure 4.23. Measured radiation pattern of the 4-element phased array. Port 1 at 8.5 GHz.

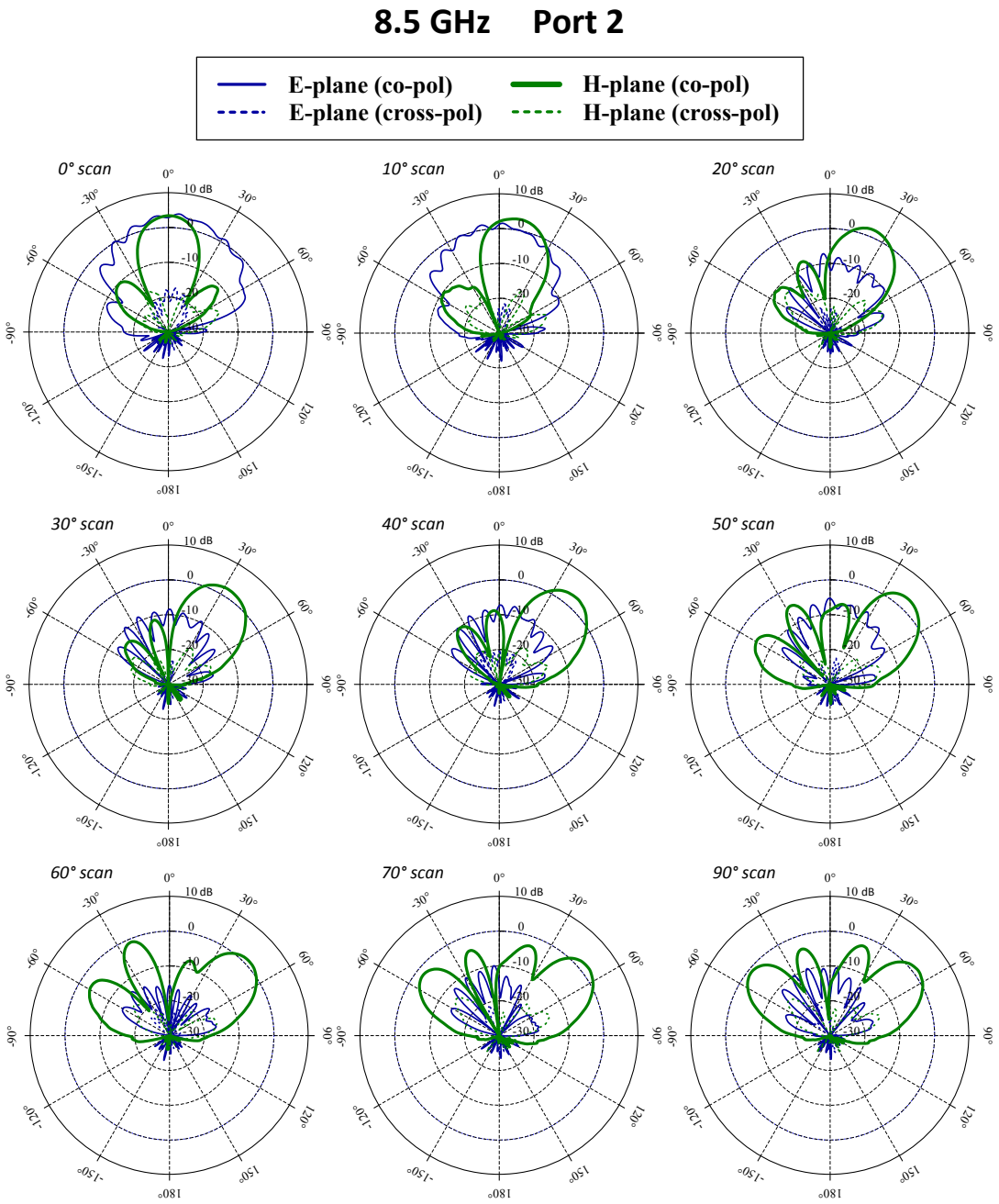


Figure 4.24. Measured radiation pattern of the 4-element phased array. Port 2 at 8.5 GHz.

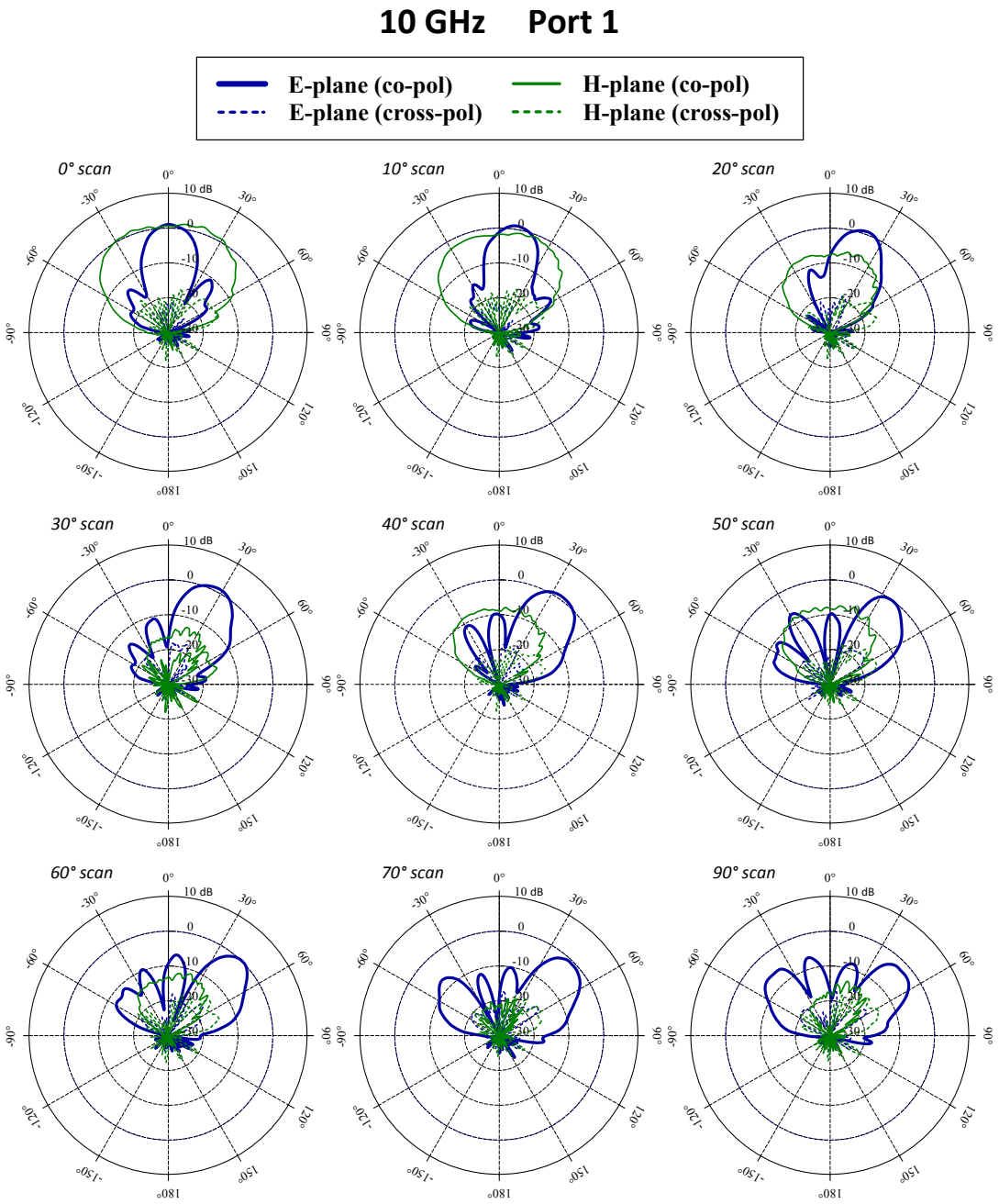


Figure 4.25. Measured radiation pattern of the 4-element phased array. Port 1 at 10 GHz.

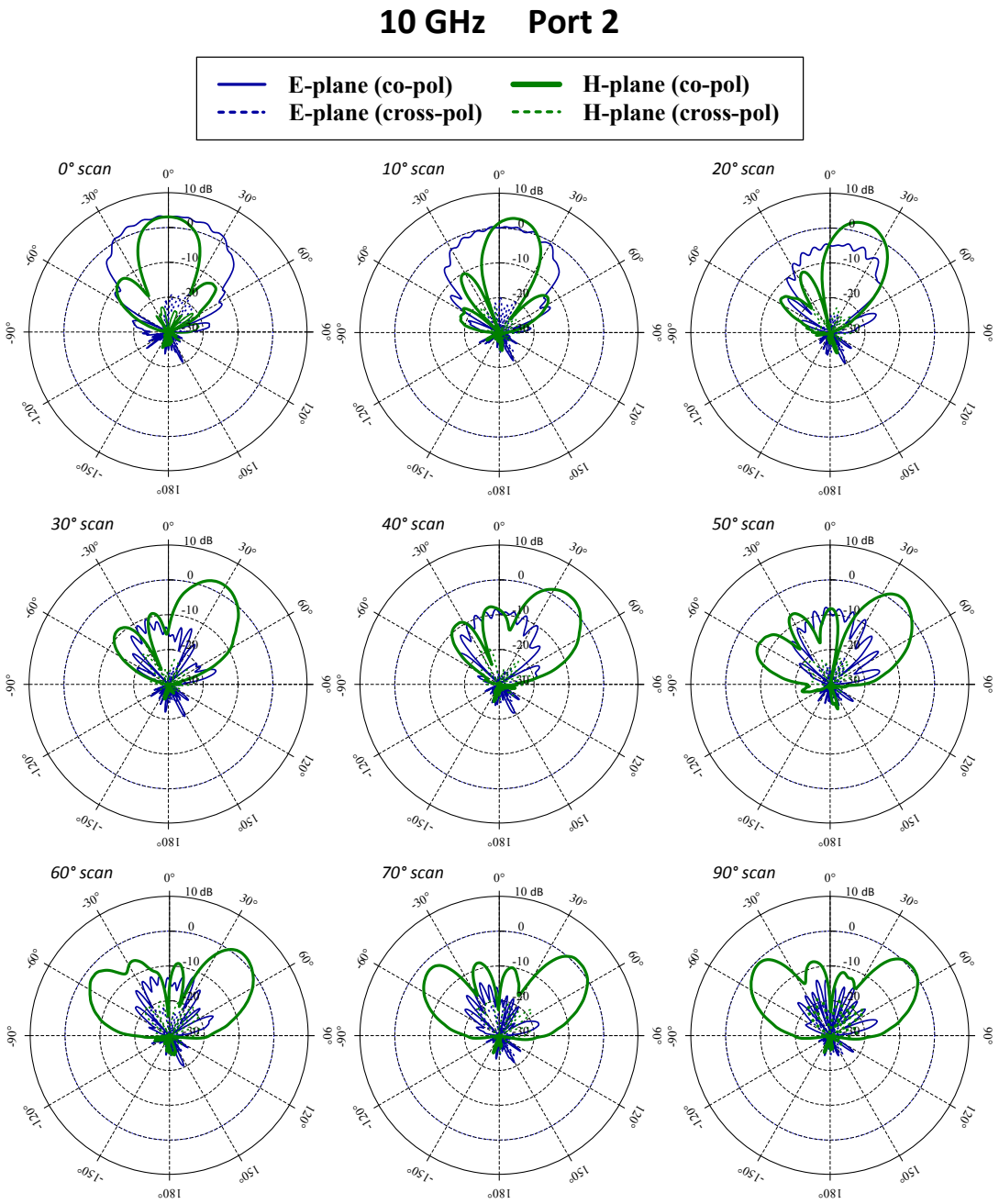


Figure 4.26. Measured radiation pattern of the 4-element phased array. Port 2 at 10 GHz.

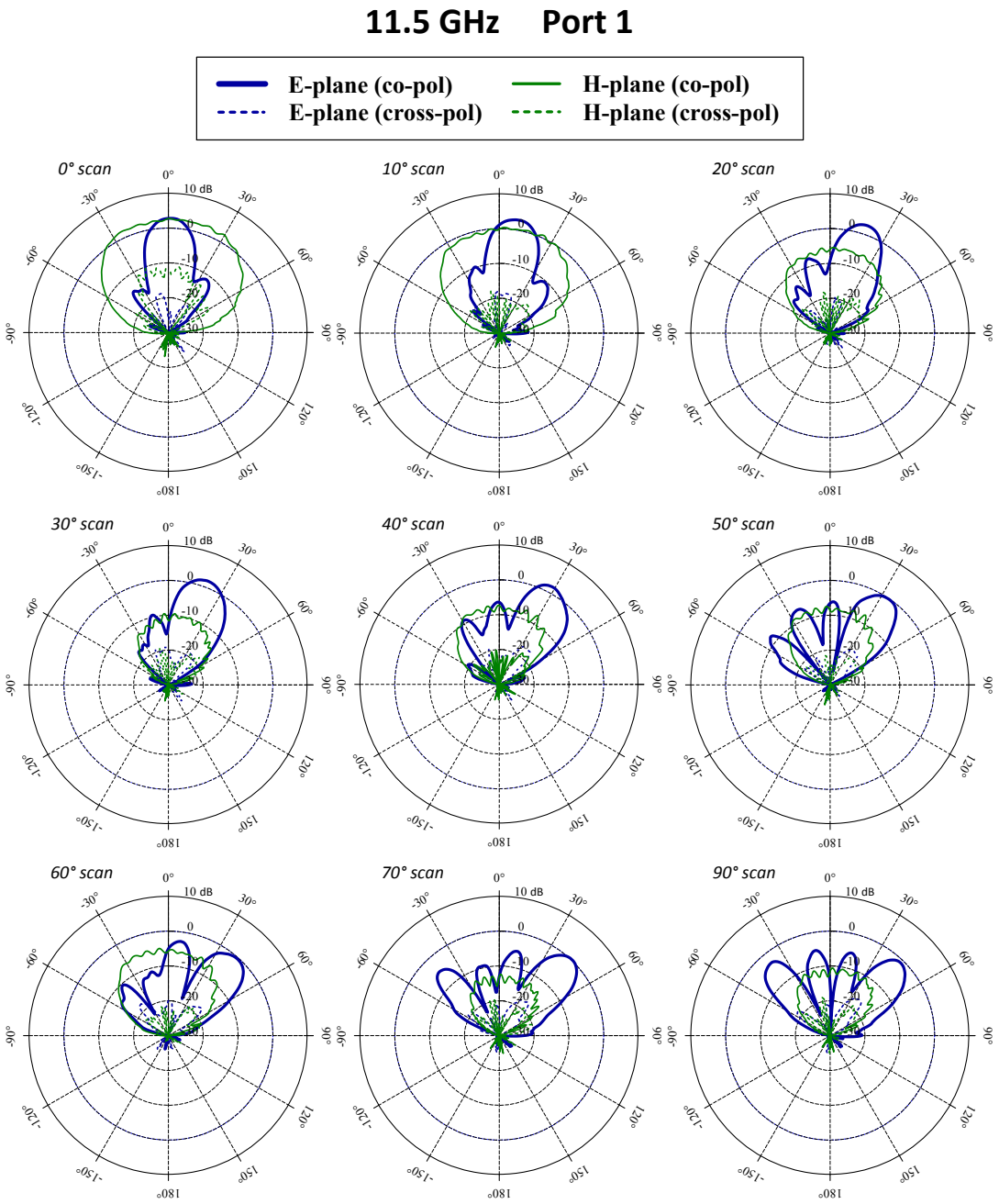


Figure 4.27. Measured radiation pattern of the 4-element phased array. Port 1 at 11.5 GHz.

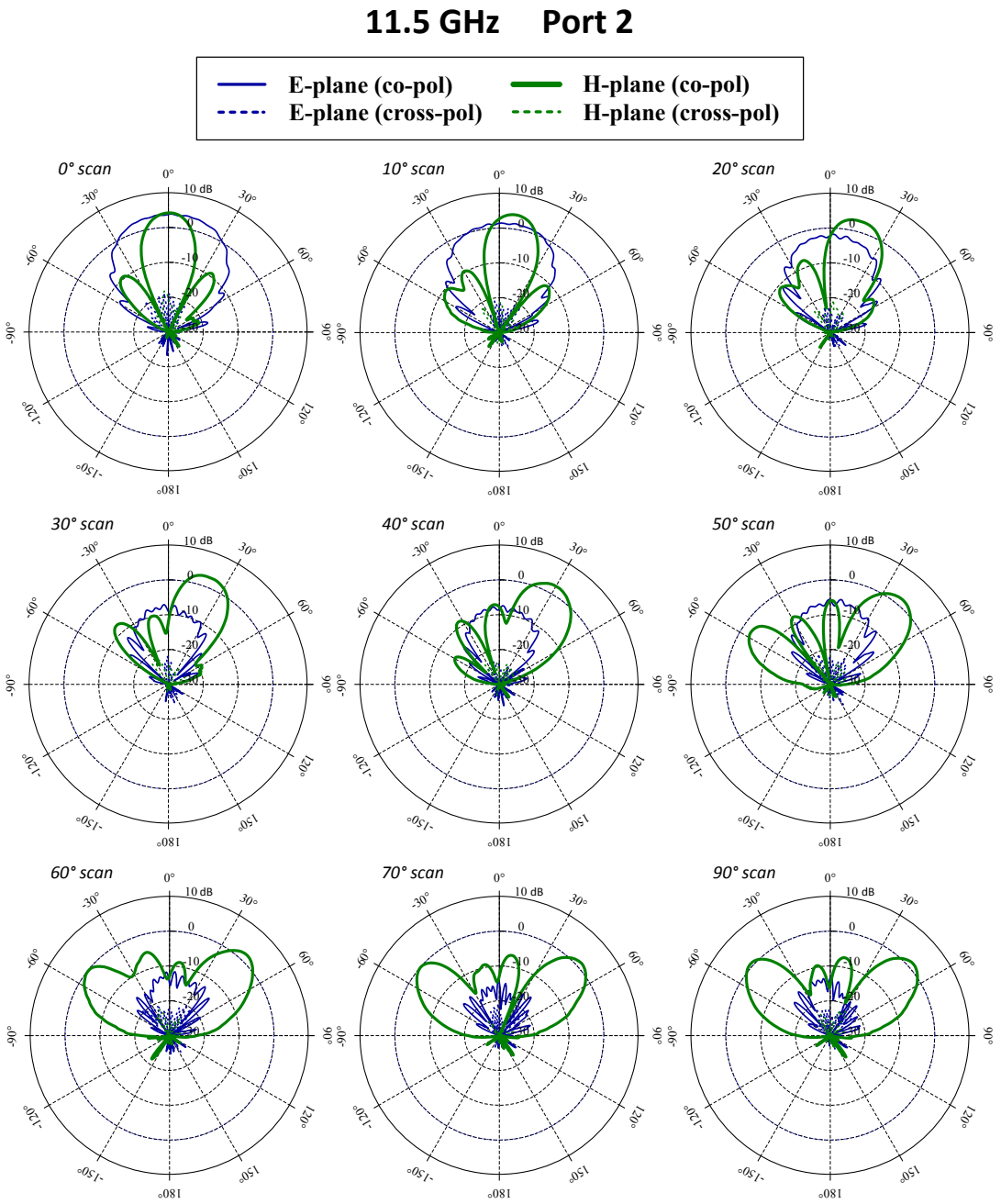


Figure 4.28. Measured radiation pattern of the 4-element phased array. Port 2 at 11.5 GHz.

Since this is a dual-polarized array, the E-plane is the scanning plane for port 1, and the H-plane is the scanning plane for port 2. At the center frequency (10 GHz), the array scans from a setting of 0° to 40° with virtually no change in peak gain (i.e., no scan loss). At the 50° setting, peak gain starts to decrease and continues through the 90° setting. This is caused by the directionality of the single-element antenna which is used. Also, the array steers to an angle slightly less than the setting would indicate. Again, this is due to the directionality of the single-element. The array factor points to the angle specified by the setting, but when it is multiplied by the single-element pattern, the 4-element array does not have an array factor with a narrow enough beamwidth to overcome the decrease in single-element gain.

The maximum scan angle of this array is approximately 45°. There are two sidelobes; the third lobe to appear is a grating lobe which becomes noticeable at a 50° scan, and at a 70° scan it dwarfs the sidelobes and approaches the magnitude of the main lobe. The sidelobes are around 10 dB below the main beam, and the cross-polarization levels are 15 - 20 dB below the main beam.

Port 1 has a measured HPBW (half-power beamwidth) of 21°-29° for the E-plane and 80°-89° for the H-plane. Port 2 has a measured HPBW of 23°-29° for the H-plane and 58°-67° for the E-plane. Note that the E-plane is the scanning plane for port 1, and the H-plane is the scanning plane for port 2. Since the single-element E-plane has a narrower beamwidth than its H-plane, it can be expected that for the array, the scanning plane of port 1 will have a narrower beamwidth than port 2, and the non-scanning plane

of port 2 will have a narrower beamwidth than port 1. This is seen in the measured HPBW.

At frequencies lower than 10 GHz, the element spacing d is less than 0.5λ. It can be seen from (4.6) that if β is constant and d decreases, then θ_0 must increase. Therefore, when β is set for a 10 GHz setting, frequencies lower than this will steer to a larger angle. The opposite is also true; higher frequencies will steer to a smaller angle. This effect is illustrated in Figs. 4.23-4.28.

At some frequencies, the main beam can actually increase in magnitude when the scan angle increases. According to theory, this should not happen, and it is probably due to two of the effects discussed earlier: the phase-shifter's phase instability with frequency and phase-shifter insertion loss variation with phase setting and frequency. The phased array is calibrated at 10 GHz. When other frequencies are used with this calibration, the phase shift can vary as seen in Fig. 4.10. This will effectively cause the calibration to be slightly off for different frequencies. Also, the phase-shifter insertion loss varies with phase setting and frequency, so the shifter setting at larger scan angles could potentially have less insertion loss than at lower scan angles. This would cause an increase in array gain for larger scan angles. This change in insertion loss with frequency and setting is illustrated in Fig. 4.9. These effects cause the slightly unusual results at some frequencies, but the array still behaves well and steers to approximately the desired scan angle at all frequencies using only a single calibration. Of course, non-calibration frequency anomalies can be corrected by a recalibration at any desired frequency.

I. Conclusion

The single-element antenna described in a previous section was used to create a 4x1 element phased array with a 0.5λ element separation at 10 GHz. A 1-to-4 power divider was designed to divide the power evenly between the four elements, and digitally controlled phase-shifters were used to steer the array. LabVIEW and a DAQ control the phase-shifters, and logic conversion circuits are required to convert from the standard 5 V TTL/CMOS logic of the DAQ to the 0/-3V complementary logic that the phase-shifters require. The array was calibrated at 10 GHz using a far-field phase-alignment method, and measurements were taken of the array's performance. The array successfully operates over a 3 GHz bandwidth with a maximum scan angle of approximately 45°.

The phased array can be improved with the addition of more elements. This will decrease the beamwidth of the array factor which will allow larger scan angles and form a more directed beam. Using a phase-shifter with a more consistent insertion loss and phase shift would also improve the array's performance. The addition of controllable amplifiers for each element could correct the phase-shifter's variable insertion loss problem and also allow for a non-uniform amplitude distribution which could narrow the array's beamwidth.

5. RECTENNA HISTORY

In wireless power transmission (WPT) systems, microwave power is transmitted wirelessly from one location to another where it is then converted into a useable form. The rectenna (rectifying antenna) is a crucial component in this system. It consists of a receiving antenna which collects microwave power and a rectifier which converts this microwave power into useable DC power. It is called a "rectenna" from the combination of "rectifier" and "antenna". The rectenna is used in numerous applications such as space power transmission (SPT) [32], wireless sensors [33], RFID tags [34], recycling ambient microwave energy [35], and powering devices in hazardous or difficult to reach places [36]. Since power transmission by wires is extremely efficient and cost effective, wireless power transmission is only a practical replacement in applications where it would be difficult or inconvenient to route power lines from the power source to the receiver or if power needs to be quickly delivered to an area without a useable power infrastructure such as for disaster relief.

Wireless power transmission was originally demonstrated by Heinrich Hertz in 1888 when he used a half-wavelength dipole spark gap to generate and transmit power at 500 MHz. He was able to detect this transmission at the receiver with the help of parabolic reflectors at both the transmitter and receiver. To generate this power, both sides of the dipole were charged with high-voltage which arced across the dipole gap and set up an oscillation along the length of the dipole which radiated from the antenna [37]-[38].

Nikola Tesla further demonstrated wireless power transmission in 1898 when he used Tesla coils to generate power at 150 kHz to light 200 light bulbs positioned 26 miles away [37] [39]. Tesla's high power wireless power transmission endeavors continued from 1899-1910 with limited success, but his activities did bring much attention to the idea [40].

During the 1920's and 1930's, researchers from the United States [41] and Japan [42] conducted feasibility studies into wireless power transmission and concluded that current technology was not sufficiently developed to make WPT feasible [37]. The major shortcoming was that devices did not exist which could transmit large amounts of electromagnetic power in a narrow beam. This shortcoming was addressed in the late 1930's when two microwave generating devices were invented. These were the velocity-modulated beam tube (now modified and known as the klystron tube) and the microwave cavity magnetron [38] [43]. Then in 1960, the US Department of Defense contracted Raytheon to build a very high-power, high-efficiency microwave tube (Amplitron) which had an output of 400 kW of CW power at 3 GHz and an efficiency of 80 percent [44]. This filled in the high-power microwave tube technology gap [38].

Another technology not sufficiently developed and preventing complete wireless power transmission systems was the conversion of microwave power to DC power [38]. The US Air Force contracted a number of companies to study the rectification of microwave power [45]. The most important study for WPT was one at Purdue University investigating the use of semiconductor diodes as an efficient rectifier [46].

Also during this time, Raytheon supported William Brown in developing a close-spaced thermionic diode rectifier to be used for a WPT demonstration [47].

In May of 1963 William Brown and others demonstrated a complete WPT system at Raytheon's Spencer Laboratory to a large group of US DOD officials who had sponsored the development of an Amplitron. For the demonstration, a magnetron was used to generate 400 W of microwave power which was transmitted to a receiver and used to drive a DC motor attached to a fan at 100 W [38].

Motivated by Raytheon's previous successes, the US Air Force expressed interest in powering aircraft with WPT. In October of 1964, William Brown and others demonstrated the sustained flight of a tethered helicopter powered by WPT [48]. This demonstration garnered much news coverage both on TV and the newspaper, including a spot on Walter Cronkite's CBS news program. This event validated the possibility of powering larger surveillance and communication aircrafts with WPT, and it was the first time that most people had heard of wireless power transmission or rectennas [37].

The Marshall Space Flight Center (MSFC) became interested in WPT as a means to power satellites from a central space station in low-Earth orbit [40] following a demonstration given by Brown in a board room where he powered a rectenna driven propeller from a 100 W microwave beam on the other side of the room [38]. This resulted in a contract in 1970 with Raytheon to build a high efficiency WPT system demonstration. An initial system was demonstrated at the MSFC in September 1970 with an overall DC-to-DC efficiency of 26%. The next few years of the contracts included work on improving rectenna design, microwave beam launching methods, and

measurement tools which culminated in a system with an overall DC-to-DC efficiency of 48%. Important improvements included using a large transmitting antenna with a Gaussian beam and minimal sidelobes. The use of a Gaussian beam meant that the power density at the center of the rectenna array may be 50 times higher than at the edge of the rectenna array. The spacing of the rectenna elements and the DC load of each element were adjusted to maximize absorbed power and efficiency [38].

In 1968, Dr. Peter Glaser introduced a concept which would change the future development of wireless power transmission [38]; this was the concept of the solar-power satellite (SPS) or space solar power (SSP) [49]. This describes a system whereby a geosynchronous satellite comprised of solar cells collects the sun's energy and transmits this energy via microwaves to a giant array of rectennas on earth's surface. The rectennas would then convert this microwave power to DC which would supply the power grid. This was proposed to be a new form of renewable energy which could potentially replace limited and polluting fossil fuels.

The proposed solar-power satellite concept led to a technical session on SPS in 1970 at the International Microwave Power Institute at Hague, and its papers were published in a special issue of the Journal of Microwave Power [50]. NASA also funded a study in 1973 on the feasibility of a satellite solar power station led by a four company team of Arthur D. Little, Raytheon, Grumman Aerospace, and Textron [51]. This led NASA to further support microwave power transmission with oversight from both the Jet Propulsion Laboratory (JPL) and Lewis Research Center (LeRC) [52]-[53]. One of the studies led to a demonstration in the Mojave Desert at JPL's Goldstone Facility in

1975 where power was transmitted one mile with a DC output of 30 kW [53]. This was an increase in transmitted power of 50 times and an increase in distance of 100 times over what had previously been demonstrated [40]. Another study led by LeRC from 1976-1977 led to a transition from a three-plane rectenna system to a two-plane system which eventually allowed the much lighter, cheaper, and simpler thin-film etched circuits to replace the old bulky standard [38]. Following this development, almost all rectennas have been etched on thin-film using photolithography.

Rectenna conversion efficiency continued to improve, and in 1977 Brown measured the highest microwave conversion efficiency ever recorded. He used a GaAs-Pt Schottky barrier diode and an aluminum bar dipole to achieve a 90.6% conversion efficiency [54].

In 1980, Canada created the Stationary High-Altitude Relay Program (SHARP) to develop a long endurance high-altitude aircraft powered by WPT [55]. In September of 1987, a 1/8 scale SHARP prototype flew for 20 minutes at an altitude of 150 meters under the power of WPT. The aircraft had a wingspan of 4.5 meters and was powered by a 10 kW source transmitted at 2.45 GHz by a parabolic dish antenna on the ground [37].

The 1980's saw the shift of WPT leadership from the United States to Japan. This was spurred by Japan's desire to meet their increasing energy demands with limited natural resources [37]. In 1983, Japan carried out their Microwave Ionosphere Nonlinear Interaction Experiment (MINIX) [56] where 830 W was transmitted from one

rocket to another in the ionosphere. This marked the first WPT experiments in the ionosphere [37].

Wireless power transmission systems tended to operate at 2.45 GHz because of its minimal attenuation through the atmosphere, abundant technology base, and its location in the center of the industrial, scientific, and medical (ISM) band. This was until ARCO Power Technologies designed a rectenna in 1991 at 35 GHz with a 72% conversion efficiency [57]. This frequency change reduced component size and aperture area; however the components at this frequency are both expensive and inefficient. Therefore, 5.8 GHz (also in the ISM research band) is used as a compromise between these frequencies because it decreases component size and aperture area without the expense and inefficiency incurred at 35 GHz. This rise in frequency, however, does result in higher attenuation through the atmosphere in thunderstorms [58].

In the 1990's, Japan conducted the Microwave Lifted Airplane Experiment (MILAX) which involved collaboration with a number of Japanese Universities and corporations [59]. A balsa wood airplane with a 2.5 meter wingspan was flown for 400 meters at an altitude of 15 meters which lasted 40 seconds. It was powered by a phased array transmitter operating at 2.411 GHz and mounted to a van. The van followed the airplane and transmitted power to rectennas mounted on the airplane [37].

In 1993, Japan conducted the International Space Year – Microwave Power Transmission in Space (ISY-METS) experiment to demonstrate WPT in space between two rockets [60]. One rocket transmitted 800 W of power to the other rocket at 2.45

GHz. The dual-polarized rectenna was developed by McSpadden and Chang at Texas A&M University [61].

In the late 1990's, dual and circular polarization became the preferred polarization for wireless power transmission, especially SSP. Microwaves passing through the earth's atmosphere undergo Faraday rotation which can cause alignment problems with linear polarization systems and severely degrade performance. Dual and circular polarization systems do not suffer this same performance degradation since broadside alignment of transmitting and receiving antennas is arbitrary. This allows both antennas of dual and circularly polarized systems to be rotated without any change in performance [37].

NASA reassessed the SSP concept with its "Fresh Look Study" from 1995-1997 and concluded that with recent technological advances, SSP could be feasible but still remained challenging [62]. This led to an SSP Exploratory Research and Technology (SERT) study in 1999 which involved researchers from around the United States. One result from the SERT study was Strassner and Chang's 2002 design of a circularly polarized rectenna array which utilized a folded dipole antenna called a dual rhombic loop antenna (DRLA). The DRLA is circularly polarized and has a large gain of 10.7 dBi. The array has a 78% RF-to-DC conversion efficiency and operates at 5.61 GHz [63]. Also in 2002, McSpadden and Mankins estimated that SPS systems have the potential to achieve 45% DC-to-DC efficiencies [64].

A serious problem which arises in WPT systems is that of beam alignment. If the microwave beam is narrow, as would be required for a long-range, high-efficiency

system, it is difficult to keep the transmitter and receiver properly aligned. This is especially problematic with SPS systems and aircraft applications since the transmission distance is large and the atmosphere can unpredictably alter the microwave beam's path. A proposed method to solve this problem is through the use of a retrodirective array [65]-[68]. This system uses a pilot beam to align the antennas and alter the direction that the transmitting antenna array is focused. The phase of the pilot beam is acquired by receivers located on each element of the transmitting array. This phase is then compared to a reference, phase conjugated and fed back to the phase controller for that element of the transmitting array.

There are some environmental concerns about the use of an SPS system. Electromagnetic exposure is the largest concern from the public, but it has been shown that potential exposure of the general public from a 5-GW reference SPS system would be 0.1 $\mu W/cm^2$ in most places and 10 $\mu W/cm^2$ in some small grating lobe areas. This is well below accepted environmental safety levels around the world which range from 100-10,000 $\mu W/cm^2$. Radio frequency interference should be a larger environmental concern. Many wireless systems operate at 2.45 GHz, and an SPS has the potential to disrupt these communications; unblocked harmonics are also a potential interference. Frequency allocation will be very important for a successful SPS system [69]. High powered microwave transmission from space will also cause some localized heating of the ionosphere. There is some concern that this could interfere with telecommunication systems [70], perturb the earth's magnetic field, or have some other environmental impact [71].

The application of wireless power transmission currently drawing the most attention is space solar power. This has become even more true with Japan's announcement that it plans to have a MW class SSP system in place in the 2020's and a commercial 1 GW class SSP system in place in the 2030's [72]. This stems from Japan's desire for unlimited clean energy to replace its limited energy resources and reliance on foreign oil. With this SSP commitment and the numerous other WPT applications, rectennas have a bright and promising future.

6. RECTENNA OPERATION THEORY *

A. Wireless Power Transmission System

Wireless power transmission (WPT) systems and wireless communication systems are based on the same technology with the Maxwell equations governing both. The fundamental difference between the two is efficiency.

In wireless communication systems, the microwave is simply a carrier of information, so bandwidth is required to encode this information, and noise is a major concern. Efficiency is not the top priority and only becomes problematic if the received power is below the noise level. However, WPT systems do not carry any information, so a very narrow band can be used, and noise is not a problem. Efficiency, however, is of paramount importance for a practical WPT system. It is dependent on component efficiencies as well as the ability to focus the microwave beam on the rectenna.

The wireless power transmission system consists of three core system blocks as seen in Fig. 6.1. The first block is the DC-to-RF transmission block which converts the DC (or AC) source power to RF power and also radiates this power. The source of the power is dependent on the application and could be photovoltaic solar cells for use in space solar power (SSP), or it could be a power plant for terrestrial applications. The power can be converted to RF by either a solid-state semiconductor circuit (below 10 GHz) or a microwave tube such as a magnetron or a klystron [38]. The efficiency of the

* Parts of this section are reprinted, with permission, from J. Hansen and K. Chang, "Diode modeling for rectenna design," in *IEEE Antennas and Propagation Society International Symposium (APSURSI)*, pp. 1077-1080, Spokane, WA, July, 2011. Copyright © 2011, IEEE.

first block η_t is the product of the DC-to-RF conversion efficiency and the transmission efficiency.

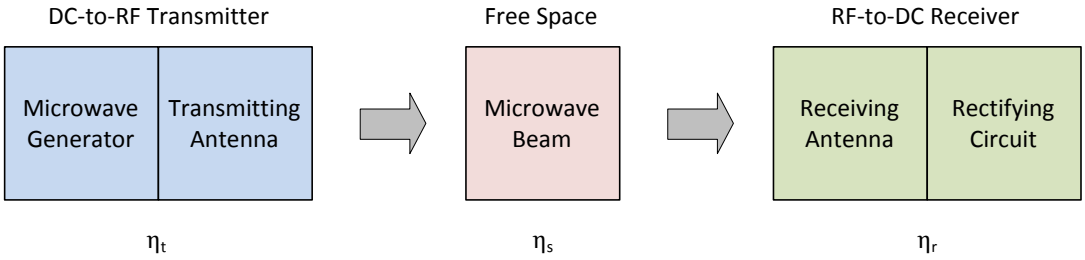


Figure 6.1. Block diagram of a wireless power transmission (WPT) system.

The second block is the free space transmission channel which accounts for propagation losses through the space separating the transmitter and receiver. Its efficiency η_s is the ratio of received power to transmitted power. For high-efficiency transmission, the microwave power must be transmitted with optimal power density. Goubau and Schwering showed that microwave power can be transmitted with efficiencies approaching 100% if the correct transmitting aperture is properly illuminated [73]-[74]. The efficiency is a function of the parameter τ [75] which is given by

$$\tau = \frac{\sqrt{A_r A_t}}{\lambda D} \tag{6.1}$$

where A_r and A_t are the aperture areas of the receiver and transmitter respectively, λ is the wavelength of the transmitted wave, and D is the distance between the transmitter and receiver. A graph of efficiency vs the parameter τ can be seen in [75]. This

presumes an aperture power density as seen in [75]. As the efficiency approaches 100%, the aperture power density approaches the shape of an edge truncated Gaussian curve. The efficiency can be approximated as

$$\eta = \left(1 - e^{-\tau^2}\right) \times 100\% \quad . \tag{6.2}$$

Larger values of τ result in higher efficiency, and the receiving and transmitting aperture areas are directly related to τ, so larger apertures result in higher transmission efficiency. Therefore, a compromise must be found between size and efficiency.

The final block is the RF-to-DC receiving block which receives the microwave power with the receiving antenna and converts this power into DC power with a rectifying circuit; this power conversion efficiency η_r determines the performance of the rectenna. This final block is called a rectenna from the combination of "rectifying" and "antenna". A block diagram of the basic components of a rectenna can be seen in Fig. 6.2 and includes an antenna, a harmonic rejection filter, a diode, a DC-pass filter (capacitor), and a load. The harmonic rejection filter allows microwaves to enter the rectifying circuit but rejects any harmonics created by the diode due to its non-linear behavior. This prevents diode harmonics from radiating back out through the antenna. The diode converts the microwave power into DC power, and the DC-pass filter reflects any unconverted microwave power back to the diode which prevents microwave power from reaching the load. By utilizing filters to block microwave power on both sides of the diode, any unconverted microwave power should be reflected back to the diode and eventually rectified to DC power, thereby improving conversion efficiency. The only

way that energy can escape from the system is in the form of DC at the output or as a reflected fundamental at the input. The DC output voltage is measured across the load. A method to further reduce harmonic radiation from the rectenna is through the use of frequency selective surfaces. In this scheme, a frequency selective surface which passes the fundamental frequency and blocks the second and third harmonics is placed in front of the rectenna. This will attenuate harmonics which attempt to radiate from the rectenna [76]-[77].

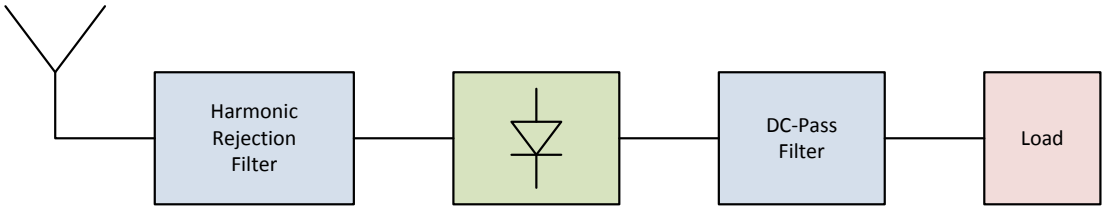


Figure 6.2. Block diagram of a rectenna.

Optimization of the rectifying circuit is important to maximize efficiency. The diode must be matched to the input power level, the input impedance, and the load. The distance between the diode and capacitor is crucial because in addition to reflecting microwave power, the capacitor is used to tune out the diode's reactance and improve conversion efficiency. This distance is typically a quarter-wavelength at the fundamental frequency, and since the capacitor can be considered a short at RF, this will have the diode seeing an open circuit at ω, 3ω, 5ω, . . . and a short circuit at 2ω, 4ω, 6ω, . . . This results in the blockage of microwave power (fundamental frequency and all

harmonics) and the passing of DC power. If the rectenna is properly designed then the efficiency becomes a property of diode parameters.

The conversion efficiency of the rectenna is calculated as

$$\eta = \frac{P_{DC}}{P_{received}} \times 100\% \tag{6.3}$$

where P_{DC} is the measured output power across the load at DC and $P_{received}$ is the RF power received by the rectenna's antenna. $P_{received}$ can be calculated using the Friis transmission equation [74] by

$$P_{received} = P_t G_t G_r \left(\lambda_0 \big/ 4\pi r \right)^2 = A_{er} \frac{P_t G_t}{4\pi r^2} \tag{6.4}$$

where the effective area of the receiving antenna is

$$A_{er} = \frac{G_r \lambda_0^2}{4\pi} \quad . \tag{6.5}$$

P_t is the transmitted power, G_t and G_r are the transmitting and receiving antenna gains respectively, λ_0 is the free space wavelength, and r is the distance between the transmitter and the rectenna. Rectenna efficiencies are oftentimes given as a function of received power density S_D which can be calculated from

$$S_D = \frac{P_{received}}{A_{er}} = \frac{P_t G_t}{4\pi r^2} \quad . \tag{6.6}$$

The overall efficiency η_{all} of the wireless power transmission system is the ratio of DC output power at the rectenna to DC (or AC) input power and is given by

$$\eta_{all} = \eta_t \times \eta_s \times \eta_r \quad . \tag{6.7}$$

B. Rectenna Operation Theory

The basic rectenna rectifying circuit is based on a half-wave rectifier with a capacitor added in shunt. It is therefore important to understand the operation of a half-wave rectifier so that the rectenna's operation can be understood. An explanation of a half-wave rectifier can be found in [78] and rectenna operation theory is discussed in [58],[79]-[81]. Some of this information will be summarized here.

A half-wave rectifier and its voltage waveforms are illustrated in Fig. 6.3. Assuming an ideal diode and a perfect input sinusoid v_s, during the positive half-cycle ($v_s \geq 0$) the diode will not conduct which will cause all of the current to flow through the load R_L. During the negative half-cycle ($v_s \leq 0$), the current will flow through the diode,

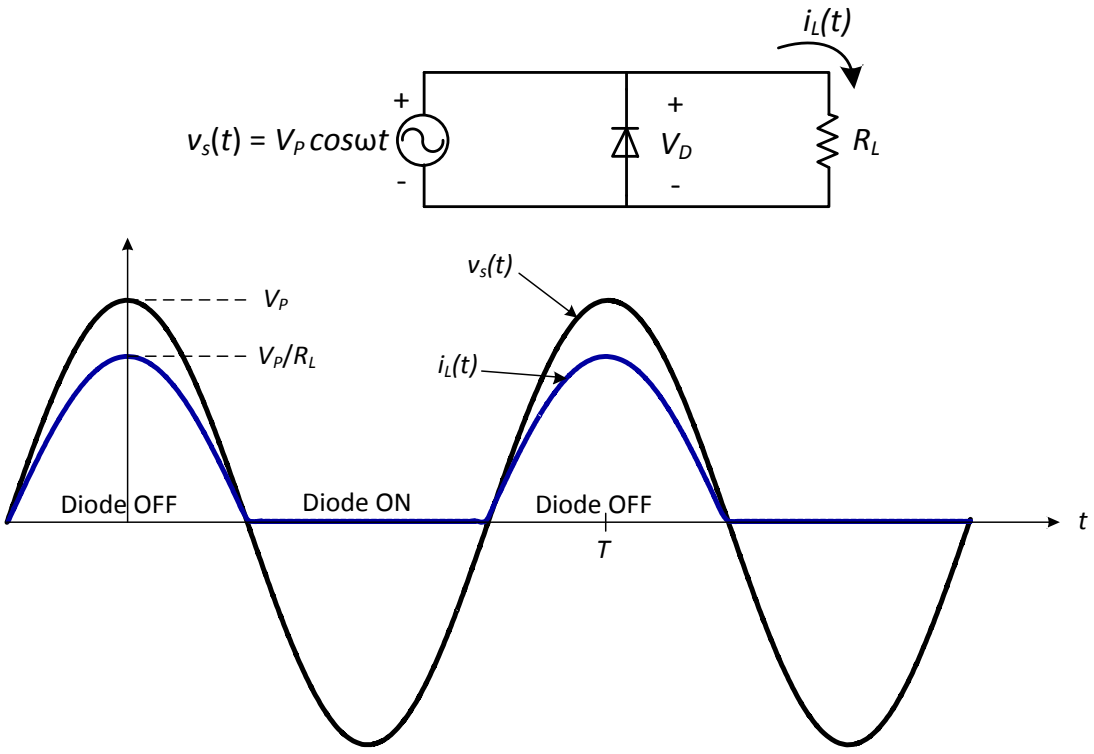


Figure 6.3. Half-wave rectifier circuit and waveforms.

bypassing the load. This will cause the current through R_L to be unidirectional and have a finite average value or a DC component. This is a rectification of the source current which can be used to generate DC from AC. The average rectified DC current $I_{L,DC}$ can be found from

$$I_{L,DC} = \frac{1}{T}\int_{T} i_L(t)\,dt = \frac{1}{T}\left(\int_{-T/4}^{T/4} \frac{V_P}{R_L}\cos\omega t\,dt\right) \tag{6.8}$$

where T is the period of the incident sinusoid and $\omega = 2\pi/T$. Solving this integral gives an average DC current of

$$I_{L,DC} = \frac{V_P}{\pi R_L}\ . \tag{6.9}$$

Applying Ohm's law gives an average rms DC voltage across the load of V_p/π.

If a capacitor (DC-pass filter) is added in shunt to this half-wave rectifier circuit, it will result in the waveform seen in Fig. 6.4. The voltage across the load is

$$v_L(t) = V_P\, e^{-t/R_L C} \tag{6.10}$$

where R_LC is the time constant and t is the time measured from the peak of the waveform where $v_L(t) = V_p$. The exponential decrease in the load voltage can be approximated by a straight line since the period is very short compared to the time constant at microwave frequencies. Equation (6.10) can be expanded using a power series [82] as

$$v_L(t) = V_P\, e^{-t/R_L C} = V_P\left[1 + \frac{-t/R_L C}{1!} + \frac{\left(-t/R_L C\right)^2}{2!} + \frac{\left(-t/R_L C\right)^3}{3!} + \ldots\right]. \qquad (6.11)$$

Because of the straight line approximation for the exponential decay, the linear terms of (6.11) are kept while the rest are discarded. The minimum value of v_L is at $t = T$ and is then

$$V_{\min} = v_L(t = T) \approx V_P\left(1 - \frac{T}{R_L C}\right) = V_P\left(1 - \frac{1}{f R_L C}\right) \ . \qquad (6.12)$$

Once the voltage $v_L(t)$ deceases to V_{min}, the diode will be non-conducting and the source voltage will be rising above V_{min} which will again raise the load voltage v_L to its peak value of V_P. The peak-to-peak ripple voltage of V_L is

$$V_r = V_{\max} - V_{\min} = \frac{V_P}{f R_L C} \ , \qquad (6.13)$$

and the average rms DC voltage across the load is

$$V_{L,DC} = \frac{V_{\max} + V_{\min}}{2} = V_P\left(1 - \frac{1}{2 f R_L C}\right) \approx V_P \ . \qquad (6.14)$$

At microwave frequencies, the time constant $R_L C$ is much larger than the period T which gives rise to a small ripple voltage V_r, resulting in an average DC load voltage of approximately V_P. Consequently, it has been shown that by adding the shunt capacitor, the average DC load voltage is raised from V_P/π to approximately V_P.

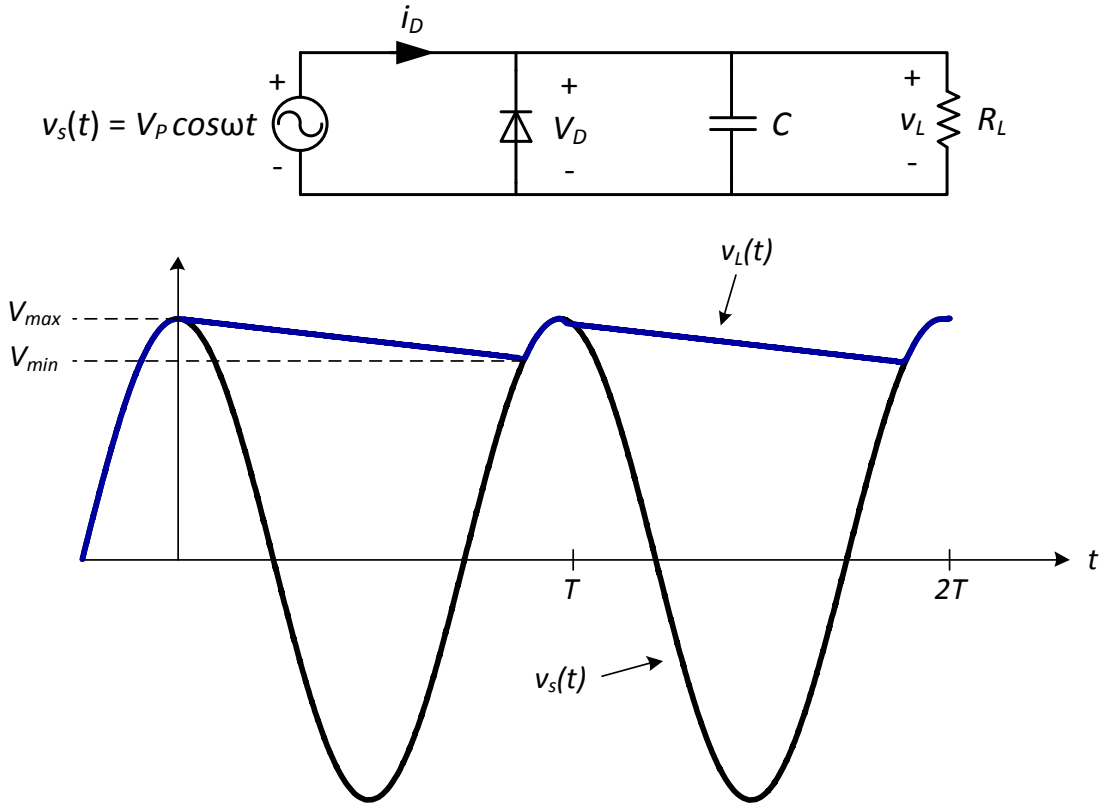


Figure 6.4. Half-wave rectifier with capacitor circuit and waveforms.

C. Analytical Model of Rectenna Efficiency

A simplified mathematical model described by Yoo [79]-[80] can be used to predict the conversion efficiency of the rectenna. This model makes a number of simplifications which prevent it from being accurate enough to be useful for precise rectenna design, but it does demonstrate rectenna behavior and reveal the effect of parameter variation on performance. It is only dependent on the simplified diode parameters and microwave circuit losses at the fundamental frequency. This model does not include harmonics created by the diode, and diode parasitics are excluded. It also assumes ideal diode characteristics (Fig. 6.5) such as: $R_j = 0$ for forward bias, $R_j = \infty$ for reverse bias, and no current through C_j during forward bias. Due to ideal condition assumptions, the results from this model can be considered as the maximum conversion efficiencies that would result from a diode with the same parameters.

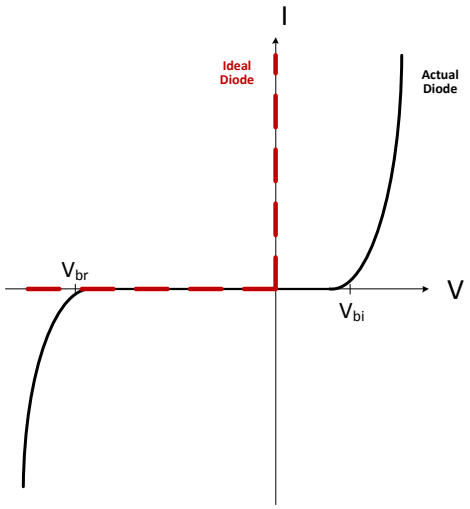


Figure 6.5. Voltage-Current curves of an ideal diode and an actual diode. The built-in turn-on voltage V_{bi} is the voltage at which forward current starts to flow through the diode. The reverse breakdown voltage V_{br} is the minimum reverse voltage which will cause reverse current flow through the diode.

The equivalent circuit of the diode rectifier which is used to derive the mathematical model is shown in Fig. 6.6. The nonlinear components R_j and C_j represent the diode junction with ideal diode conditions applied: junction resistance $R_j = 0$ for forward bias and $R_j = \infty$ for reverse bias. The voltage V_D is the voltage across the diode and the voltage V_j represents the voltage across the diode's junction. The diode's voltage waveforms are displayed in Fig. 6.7. The angle θ_{on} is the forward bias turn-on angle and ϕ is the phase lag of the junction voltage behind the incident diode voltage.

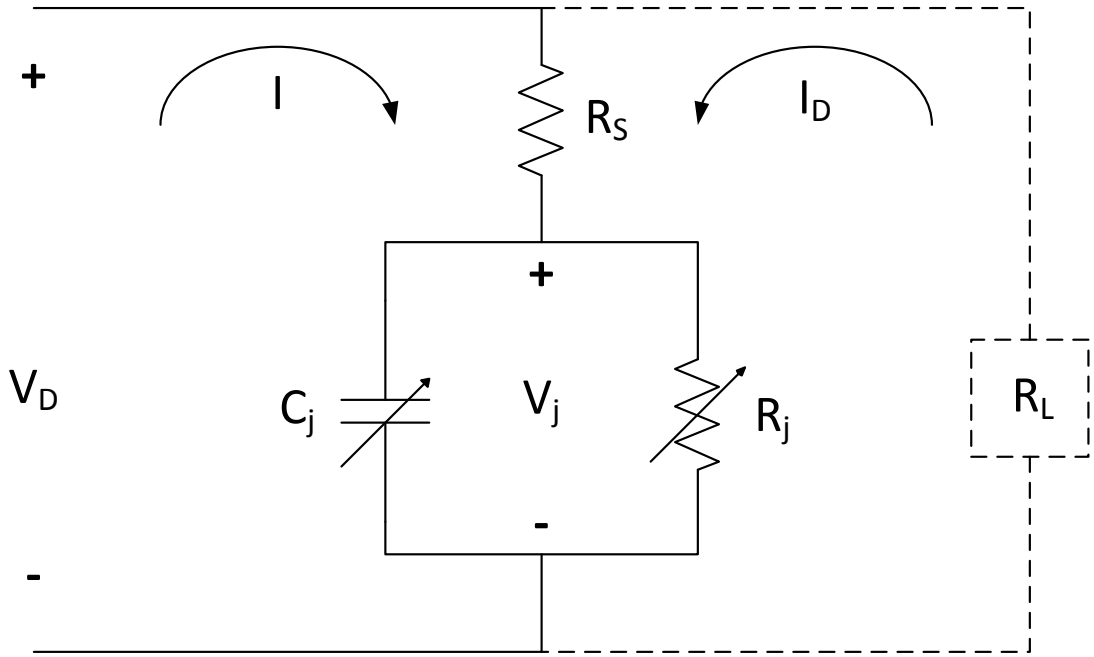


Figure 6.6. Equivalent circuit of the diode rectifier.

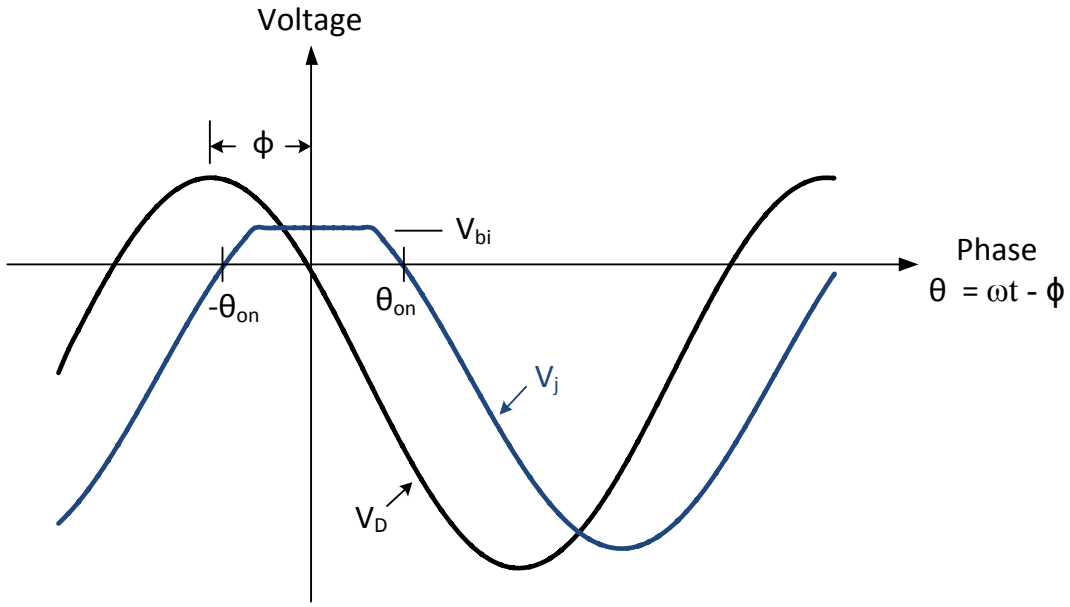


Figure 6.7. Voltage waveforms of the diode rectifier. V_D is the voltage across the physical diode, and V_j is the voltage across the diode junction.

The incident diode voltage V_D is composed of a fundamental frequency and a DC component. It can be expressed as

$$V_D = -V_0 + V_1 \cos(\omega t) \tag{6.15}$$

where V_0 is the self-biasing DC output voltage across the load R_L and V_1 is the peak voltage of the incident microwave signal. The incident microwave signal produces a DC output voltage which biases the diode. This biasing level increases through a few cycles and also as the input power level increases. The diode junction voltage is

$$V_j = \begin{cases} -V_{j0} + V_{j1}\cos(\omega t - \phi) & \text{, when diode is off} \\ V_{bi} & \text{, when diode is on} \end{cases} \tag{6.16}$$

where V_{j0} is the DC component of the diode junction voltage and V_{j1} is the peak voltage of the first harmonic of the diode junction voltage when the diode is off ($V_j < V_{bi}$). When the diode is on, the junction voltage V_j is equal to the built-in turn-on voltage of the diode V_{bi} which is set to zero in this derivation (ideal diode). The phase ϕ is the angle that the diode's junction voltage lags the incident microwave's voltage.

The DC output voltage V_0 can be determined from the junction voltage V_j. The average value of V_j in each cycle can be calculated as

$$V_{j,dc} = \frac{1}{2\pi}\int_{0}^{2\pi} V_j \, d\theta = \frac{1}{2\pi}\int_{-\theta_{on}}^{\theta_{on}} V_{bi} \, d\theta + \frac{1}{2\pi}\int_{\theta_{on}}^{2\pi-\theta_{on}} (-V_{j0} + V_{j1}\cos\theta)\, d\theta \,. \qquad (6.17)$$

The first term in (6.17) corresponds to the forward-biased region, and the second term corresponds to the reverse-biased region. Note that the phase lag ϕ is assumed to be zero for this derivation. Integrating (6.17) gives

$$V_{j,dc} = V_{bi}\frac{\theta_{on}}{\pi} - V_{j0}\left(1 - \frac{\theta_{on}}{\pi}\right) - \frac{V_{j1}}{\pi}\sin\theta_{on} \quad . \qquad (6.18)$$

Applying Kirchhoff's voltage law along the DC path as illustrated in Fig. 6.6 by the dotted line gives

$$V_0 = V_{j,dc}\frac{R_L}{R_S + R_L} \,. \qquad (6.19)$$

The diode switching occurs when the junction voltage V_j is equal to the turn-on voltage V_{bi} and $\omega t = \theta_{on}$, so from (6.16),

$$V_{bi} = V_j = -V_{jo} + V_{j1}\cos\theta_{on} \qquad (6.20)$$

which can be rearranged to give an expression for the turn-on angle

$$\cos\theta_{on} = \frac{V_{bi} + V_{j0}}{V_{j1}} \quad . \tag{6.21}$$

When the diode is off, R_j is infinite and Kirchhoff's voltage law can be applied to the RF loop in Fig. 6.6 to give

$$-V_D + IR_S + V_j = 0\,, \tag{6.22}$$

and since all current is flowing through C_j when the diode is off,

$$I = \frac{d(C_j V_j)}{dt}\,. \tag{6.23}$$

Equations (6.22) and (6.23) can be combined to give

$$I = \frac{d(C_j V_j)}{dt} = \frac{V_D - V_j}{R_S} \tag{6.24}$$

where C_j can be expressed as a harmonic function of V_D:

$$C_j = C_0 + C_1\cos(\omega t - \phi) + C_2\cos(2\omega t - 2\phi) + ... \tag{6.25}$$

Substituting (6.15), (6.16), and (6.25) into (6.24) and neglecting terms higher than the second harmonic gives

$$\omega R_S\left(C_1 V_{j0} - C_0 V_{j1}\right)\sin\theta = V_{j0} - V_0 + \left(V_1\cos\phi - V_{j1}\right)\cos\theta - V_1\sin\phi\sin\theta \tag{6.26}$$

where $\theta = \omega t - \phi$ for simplicity. Since this equation is valid whenever the diode is off, each sinusoidal term should separately be satisfied during the diode's off period as follows:

$$V_{j0} = V_0 \tag{6.27}$$

$$V_{j1} = V_1\cos\phi \tag{6.28}$$

$$V_1 \sin\phi = \omega R_S \left(C_0 V_{j1} - C_1 V_{j0} \right) \ . \tag{6.29}$$

An expression for the turn-on angle can be derived by combining (6.18), (6.19), (6.21), and (6.27) into

$$\tan\theta_{on} - \theta_{on} = \frac{\pi R_S}{R_L \left(1 + \frac{V_{bi}}{V_0} \right)} \ . \tag{6.30}$$

It can be seen that the ratio of R_s to R_L and the ratio of V_{bi} to V_0 are the only variables that determine θ_{on}. Note that the input power will affect V_0, but the expression for θ_{on} is not directly dependent on input power.

The diode efficiency η is defined as

$$\eta = \frac{P_{DC}}{P_{Loss} + P_{DC}} \tag{6.31}$$

where P_{DC} is the DC output power and P_{Loss} is the power dissipated by the diode. P_{DC} and P_{Loss} are defined as

$$P_{DC} = \frac{V_0^2}{R_L} \tag{6.32}$$

$$P_{Loss} = Loss_{on,R_S} + Loss_{off,R_S} + Loss_{on,diode} \tag{6.33}$$

where $Loss_{on,Rs}$ and $Loss_{off,Rs}$ are the losses from R_S when the diode is on and off, respectively. $Loss_{on,diode}$ is the loss from the diode when the diode is on. The loss terms of (6.33) can be expressed as

$$Loss_{on,R_S} = \frac{1}{2\pi} \int_{-\theta_{on}}^{\theta_{on}} \frac{\left(V_D - V_{bi} \right)^2}{R_S} d\theta \tag{6.34}$$

$$Loss_{off,R_S} = \frac{1}{2\pi}\int_{\theta_{on}}^{2\pi-\theta_{on}} \frac{\left(V_D - V_j\right)^2}{R_S} d\theta \tag{6.35}$$

$$Loss_{on,diode} = \frac{1}{2\pi}\int_{-\theta_{on}}^{\theta_{on}} \frac{\left(V_D - V_{bi}\right)V_{bi}}{R_S} d\theta \quad . \tag{6.36}$$

Since the junction resistance is assumed to be infinite during the off cycle of the diode, the loss through the diode during the off cycle is neglected. These power loss terms are the time-average products of the voltage across each element and the current through this element. The total power lost from the series resistance R_s can be found from using (6.15), (6.16), (6.24), and (6.28) in (6.34) and (6.35) and setting $\phi = 0$ as

$$L_{R_S} = \frac{1}{2\pi R_S}\left[\int_{-\theta_{on}}^{\theta_{on}} \left(-V_0 - V_{bi} + V_1\cos\theta\right)^2 d\theta + \left(\omega R_S C_j V_1\right)^2 \int_{\theta_{on}}^{2\pi-\theta_{on}} \sin^2\theta\, d\theta\right], \tag{6.37}$$

and the power lost from the diode junction can be found from substituting (6.15) into (6.36) and setting $\phi = 0$ as

$$L_{diode} = \frac{1}{2\pi R_S}\int_{-\theta_{on}}^{\theta_{on}} V_{bi}(-V_0 - V_{bi} + V_1\cos\theta)d\theta \quad . \tag{6.38}$$

During the diode off cycle, the voltage drop across R_s is so small that the phase difference ϕ can be set to zero in (6.28) to obtain

$$V_{j1} = V_1 \quad . \tag{6.39}$$

V_1 in (6.38) can be determined when the diode is off by substituting (6.16), (6.27), and (6.39) into (6.20) as

$$V_1 = \frac{V_0 + V_{bi}}{\cos\theta_{on}} \quad . \tag{6.40}$$

Finally, an expression for the diode efficiency η can be found by using (6.32), (6.33), (6.37), and (6.38) in (6.31). This gives

$$\eta_D = \frac{1}{1 + A + B + C} \tag{6.41}$$

where

$$A = \frac{R_L}{\pi R_S}\left(1 + \frac{V_{bi}}{V_0}\right)^2 \left[\theta_{on}\left(1 + \frac{1}{2\cos^2\theta_{on}}\right) - \frac{3}{2}\tan\theta_{on}\right] \tag{6.42}$$

$$B = \frac{R_S R_L C_j^2 \omega^2}{2\pi}\left(1 + \frac{V_{bi}}{V_0}\right)\left(\frac{\pi - \theta_{on}}{\cos^2\theta_{on}} + \tan\theta_{on}\right) \tag{6.43}$$

$$C = \frac{R_L}{\pi R_S}\left(1 + \frac{V_{bi}}{V_0}\right)\frac{V_{bi}}{V_0}\left(\tan\theta_{on} - \theta_{on}\right). \tag{6.44}$$

The input impedance of the diode can be found from the current I which flows through R_S during one cycle by utilizing a Fourier series as

$$I = I_0 + I_{1r}\cos(\omega t) + I_{1i}\sin(\omega t) \tag{6.45}$$

where I_0 is the DC component and I_{1r} and I_{1i} are the real and imaginary parts of the fundamental frequency component, respectively. Using (6.24) and $\theta = \omega t - \phi$, these current components are given by

$$I_0 = \frac{1}{2\pi R_S}\left[\int_{-\theta_{on}}^{\theta_{on}} \left(V_D - V_{bi}\right) d\theta + \int_{\theta_{on}}^{2\pi - \theta_{on}} \left(V_D - V_j\right) d\theta\right] \tag{6.46}$$

$$I_{1r} = \frac{1}{\pi R_S}\left[\int_{-\theta_{on}}^{\theta_{on}} \left(V_D - V_{bi}\right)\cos(\theta + \phi) d\theta + \int_{\theta_{on}}^{2\pi - \theta_{on}} \left(V_D - V_j\right)\cos(\theta + \phi) d\theta\right] \tag{6.47}$$

$$I_{1i} = \frac{1}{\pi R_S}\left[\int_{-\theta_{on}}^{\theta_{on}} \left(V_D - V_{bi}\right)\sin(\theta+\phi)d\theta + \int_{\theta_{on}}^{2\pi-\theta_{on}} \left(V_D - V_j\right)\sin(\theta+\phi)d\theta\right], \quad (6.48)$$

and the input impedance of the diode at the fundamental frequency expressed as a phasor is

$$Z_D = \frac{V_1}{I_{1r} - jI_{1i}} \quad . \quad (6.49)$$

Assuming that no current flows through C_j when the diode is on and that all of the current flows through C_j when the diode is off, the diode current in one cycle can be found from

$$I_{1r} - jI_{1i} = \frac{1}{\pi R_S}\int_{-\theta_{on}}^{\theta_{on}} \left(-V_0 - V_{bi} + V_1\cos\theta\right)\cos\theta\, d\theta + j\frac{\omega C_j V_1}{\pi}\int_{\theta_{on}}^{2\pi-\theta_{on}} \sin^2\theta\, d\theta \,. \quad (6.50)$$

The input impedance of the diode can then be written as

$$Z_D = \frac{\pi R_S}{\cos\theta_{on}\left(\frac{\theta_{on}}{\cos\theta_{on}} - \sin\theta_{on}\right) + j\omega R_S C_j\left(\frac{\pi-\theta_{on}}{\cos\theta_{on}} + \sin\theta_{on}\right)} \quad . \quad (6.51)$$

If impedance matching is utilized to tune out the reactance of the diode impedance, the diode input impedance can be simplified to

$$R_D = \frac{1}{\text{Real}(Y_D)} = \frac{\pi R_S}{\cos\theta_{on}\left(\frac{\theta_{on}}{\cos\theta_{on}} - \sin\theta_{on}\right)} \quad . \quad (6.52)$$

Equation (6.30) can be used to find the turn-on angle θ_{on}, and once this is known, (6.41)-(6.44) and (6.52) can be used to calculate the theoretical maximum conversion efficiency and input resistance of a diode as a function of fundamental diode parameters. As an

example, a plot of conversion efficiency versus a normalized parameter $\omega R_S C_j$ for different R_S/R_L ratios is displayed in Fig. 6.8. Also, plots of conversion efficiency and input resistance versus load resistance for a diode with parameters $C_j = 2$ pF and $R_S = 0.5$ Ω operating at 5.8 GHz are shown in Figs. 6.9-6.10.

The junction capacitance C_j can be calculated from

$$C_j = C_{j0}\sqrt{\frac{V_{bi}}{V_{bi} + |V_0|}} \tag{6.53}$$

where C_{j0} is the zero bias junction capacitance.

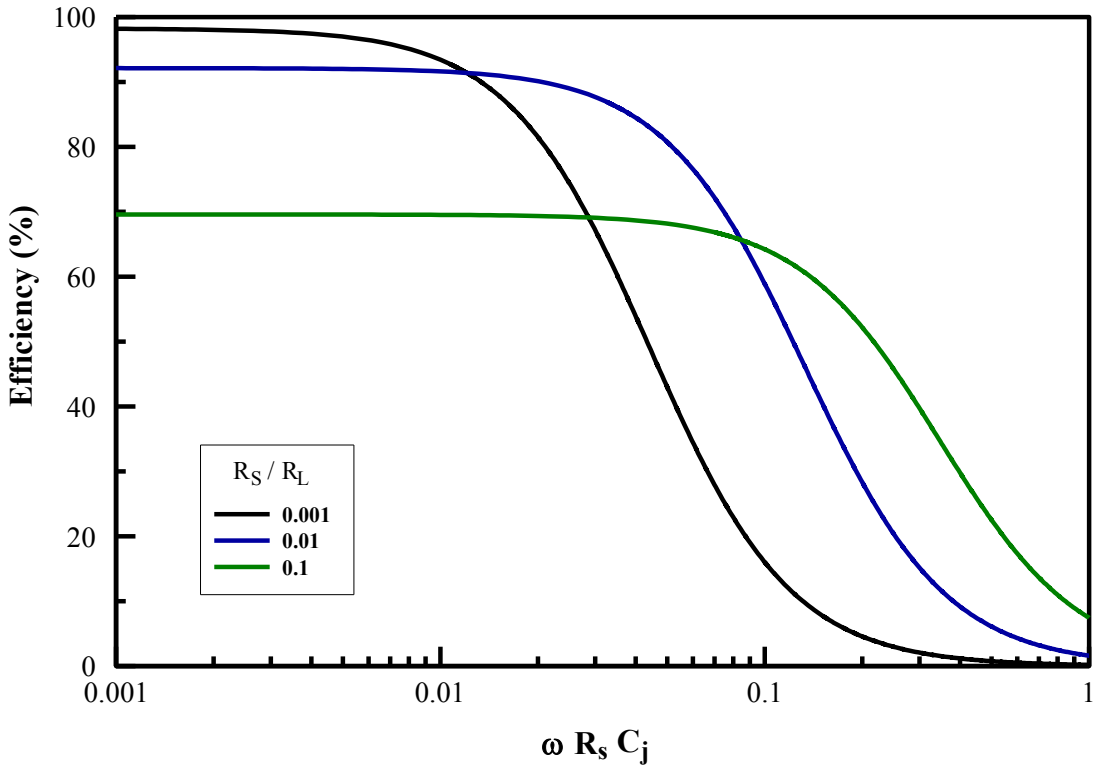


Figure 6.8. Conversion efficiency of an ideal diode versus the normalized parameter $\omega R_S C_j$ for different R_S/R_L ratios.

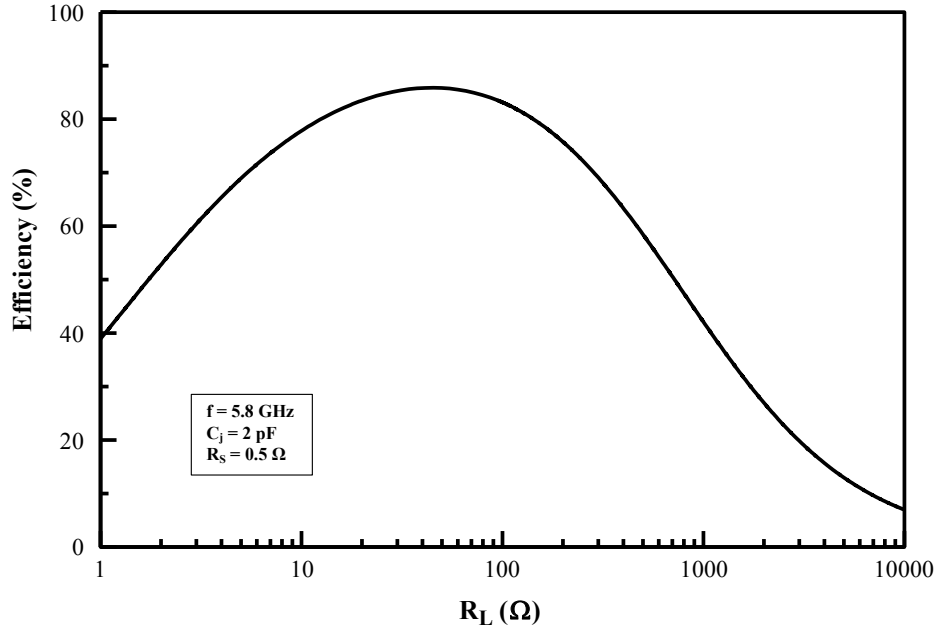


Figure 6.9. Conversion efficiency of an ideal diode versus load resistance (for a 5.8 GHz ideal diode with set parameters).

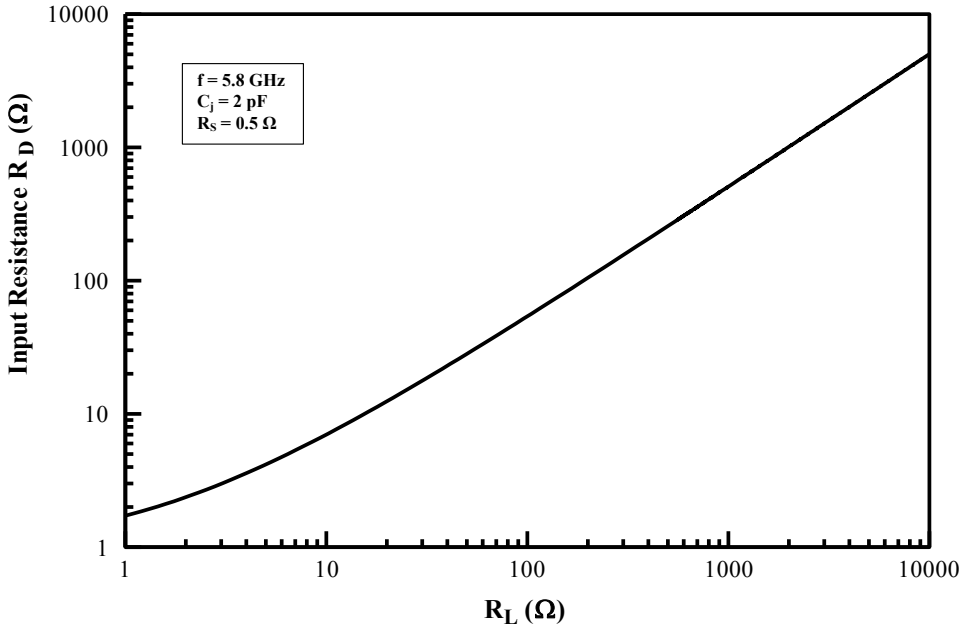


Figure 6.10. Input resistance of an ideal diode versus load resistance (for a 5.8 GHz ideal diode with set parameters).

D. Effect of Input Power on Rectenna Efficiency

The magnitude of the input power to the rectenna has a dramatic effect on its power conversion efficiency. With increasing input power, the diode conversion efficiency will rise to a point and then start to decrease [79].

At low input power levels, the conversion efficiency is low because the diode voltage is below or comparable to the turn-on voltage V_{bi} which means that the diode is not fully turned on. As the diode voltage exceeds the turn-on voltage, this effect disappears and the conversion efficiency levels off. Also, as power increases, the higher order harmonics will come into play and decrease efficiency. Finally, efficiency abruptly decreases when the power level increases to a point where the diode voltage swing exceeds the diode's breakdown voltage V_{br}. The critical point where the breakdown voltage effect takes over is when the power level is equal to $V^2_{br}/4R_L$. These effects combine to give a net conversion efficiency curve [79].

E. Diode Selection

The diode is the most crucial component in rectenna design because in a well-designed rectenna, it has the largest effect on efficiency. GaAs and Si diodes can both be used in a rectenna. GaAs is usually preferable for high efficiency because it has more than six times the electron mobility of Si, but Si has higher thermal conductivity which is better for long-term reliability.

Diode selection is dependent on system design parameters such as frequency and input power. In general, a small series resistance R_S and low turn-on voltage V_{bi} are

preferred because this will reduce rectenna losses. The diode's reverse breakdown voltage V_{br} limits the diode's power handling capacity and should be large enough to handle the expected power level. V_{br} and the DC output voltage V_0 should be chosen such that $V_0 < V_{br}/2$. Since all parameters are related to each other, diode parameter tradeoffs must be made.

7. DIODE MODEL *

A. Background and Introduction

Computer simulation has become an increasingly popular design tool for microwave circuits and systems, and the design of a rectenna is no exception. The capability of simulating a rectenna design will allow for the analysis of new ideas and tuning of parameters without the added expense and time of fabricating and measuring each new design to test its validity.

All parts of the rectenna, with the exception of the diode, can be easily modeled using standard microwave simulation software. The antenna and transmission lines can be modeled using EM modeling software such as IE3D [12] or HFSS [13], and the capacitor and load (resistor) are standard components which can be modeled in any software capable of circuit-level simulation, such as Agilent ADS [83]. However, diodes present a problem with off-the-shelf modeling because the parameters of different diodes vary from each other, so there is not a uniform standard model which works for all diodes. Models of a few popular diodes are usually included in circuit level simulator software or may be provided by the manufacturer, but if a diode, which is not provided with a model, is to be used in a rectenna then a model for this particular diode should be created. After this diode model is created, it will be possible to simulate the entire rectenna design and avoid unnecessary fabrication and testing.

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Device models are usually not generic, and the limitations of the model must be appreciated. The model must be designed to work in the target application. For use in rectenna modeling, a nonlinear harmonic balance simulator is utilized, so the model must be compatible with this type of simulation.

A harmonic balance analysis [84]-[85] can be used to solve the steady-state of a nonlinear circuit which is excited by a single-tone. It works in the frequency domain by balancing currents and voltages until Kirchhoff's Current Law is satisfied at the fundamental frequency and a specified number of harmonics. An initial guess is made at node voltages which are used to calculate currents. These currents are then checked for adherence to Kirchhoff's Current Law. An iterative process is used to update the node voltages, and the procedure is repeated until Kirchhoff's Current Law is satisfied.

The model should imitate the operation of the device within the required simulation environment by generating similar characteristics to the physical diode by utilizing an analogous system. Device models can be either physical device models or equivalent circuit models.

Equivalent circuit models [86] are common for circuit design applications because they are relatively easy to implement, and they can seamlessly mimic the electrical performance of the device being modeled. However, it can be difficult to relate the model parameters to the physical device parameters because of the non-linear behavior, frequency dependence, and biasing of most devices. This can restrict the use of equivalent circuit models because they are generally not suitable for predicting the performance of new devices or modeling large-signal operation [87].

Physical device models are based on the physical operation of carrier transport and therefore provide greater insight into how the device operates. They are usually not restricted to certain applications like an equivalent circuit model, and they can be used to predict the performance of new devices. However, these advantages come at the price of requiring considerably more computer simulation time and memory [87].

B. Diode Equivalent Circuit

The diode might be the simplest modern solid-state device in existence and is probably the easiest to accurately characterize and model; the model can be both simple and accurate [88]. The equivalent circuit model of a diode takes much less computing power to evaluate than a physical device model, and it is simple and accurate under a wide range of conditions. It is for these reasons that a standard diode equivalent circuit will be the basis of the rectenna's diode model. The standard diode model which will be used is illustrated in Fig 7.1 [89].

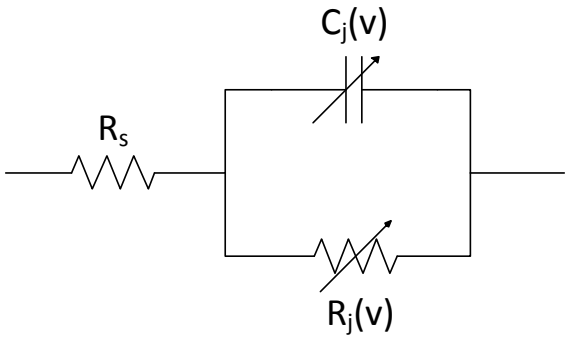


Figure 7.1. Equivalent circuit of the diode.

Mathematical expressions for the voltage-dependent parameters can be obtained from analytical expressions based on the internal physics of the diode [90]. The parasitic series resistance R_s represents the spreading resistance which is associated with the bonding wire, the semiconductor substrate, and the ohmic contact. The series resistance is also nonlinear but only varies slightly under forward bias, so it can be treated as linear [88].

The voltage-dependent resistor R_j is interchangeable with a voltage-dependent current source I_j. This nonlinear current can be expressed by a simple expression which is derived assuming that conduction occurs primarily by means of the thermionic emission of electrons over a barrier. Conduction by other means such as tunneling does occur with a limited scope, but this thermionic emission assumption is valid for Schottky diodes operating close to room temperature with moderate doping densities [88]. The nonlinear current source follows the Shockley equation:

$$I_j = I_0\left(e^{V_j/nV_t} - 1\right) \tag{7.1}$$

where V_t is the thermal voltage which is equal to kT/q and has a value of 25.69 mV at 25°C, k is Boltzmann's constant, T is absolute temperature, q is the electron charge, and V_j is the voltage across the junction. The model parameters I_0 and n are used to model the effects of the junction and differ from diode to diode. The current parameter or saturation current I_0 is a proportionality constant, and the ideality factor n accounts for imperfections in the junction [88].

The voltage-dependent capacitor C_j is the junction capacitance. In a Schottky diode, a reverse-biased junction creates a wider depletion region, and more electrons move to the anode which leaves behind more positive charge. A forward-biased junction creates a narrower depletion region, and fewer electrons move to the anode which results in less stored charge. This creates a situation where a negative voltage increases the amount of negative charge stored on the anode and a positive voltage reduces the amount of negative charge stored [88]. This results in the junction behaving as a nonlinear capacitor [89] which can be expressed as

$$C_j = \frac{C_0}{\left(1 - V_j / V_0\right)^{\gamma}} \tag{7.2}$$

where V_0 is the diffusion potential, C_0 is the zero-voltage junction capacitance, γ is the grading coefficient, and V_j is the voltage across the junction. It can be assumed that the junction is uniformly doped and the transition between regions is abrupt ($\gamma = 0.5$) [88] which simplifies the equation to

$$C_j = C_0 \sqrt{\frac{V_0}{V_0 - V_j}} \ . \tag{7.3}$$

The diode comes in a package, and at microwave frequencies the effect of this packaging must be taken into account. The packaging creates parasitics which limit the performance of the diode. A package parasitic inductance L_p is added to account for the inductance of wires or conductors used to connect the die inside to the exterior terminations. A package parasitic capacitance C_p and a beam-lead parasitic capacitance

C_b are added to account for the capacitance due to the dielectric constant of the solid packaging. The modified equivalent circuit can be seen in Fig 7.2 [89].

The values of the equivalent circuit's parameters can be calculated by measuring the diode's small-signal S-parameters. The voltage dependence can be determined by measuring the small-signal S-parameters at various bias levels and combining these measurements to create a large-signal model [91]. The procedure involves altering the values of the circuit parameters until the S-parameters of the equivalent circuit sufficiently match the measured S-parameters. An optimization procedure is generally required for this multi-variable procedure, and unique solutions usually do not exist [92]. Software packages can be used to fit the S-parameters [93], but if this software is not available, another method must be used.

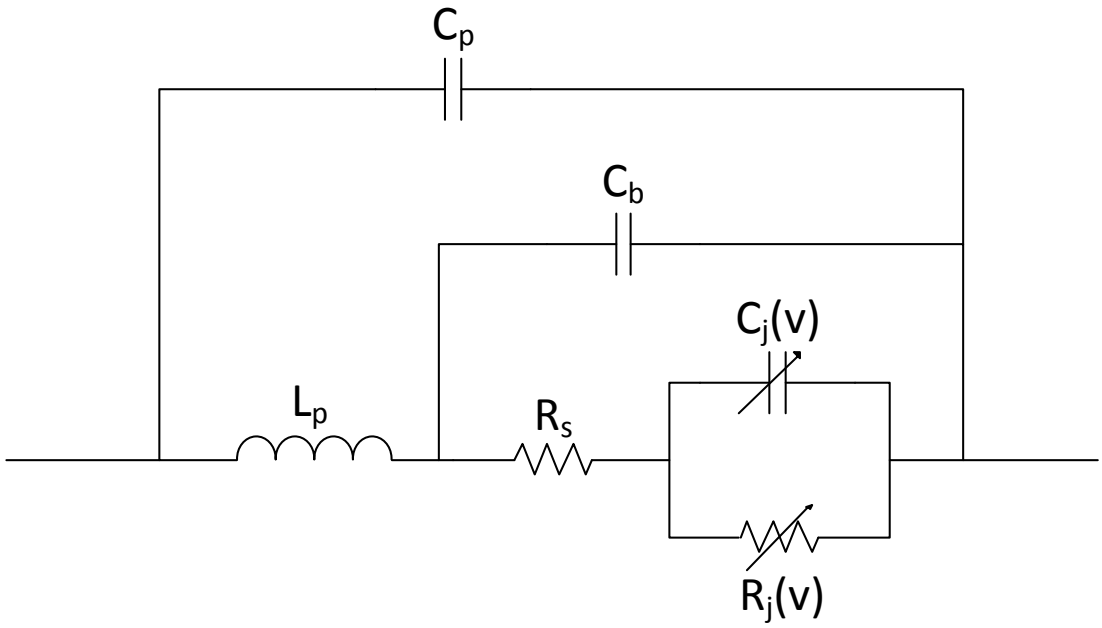
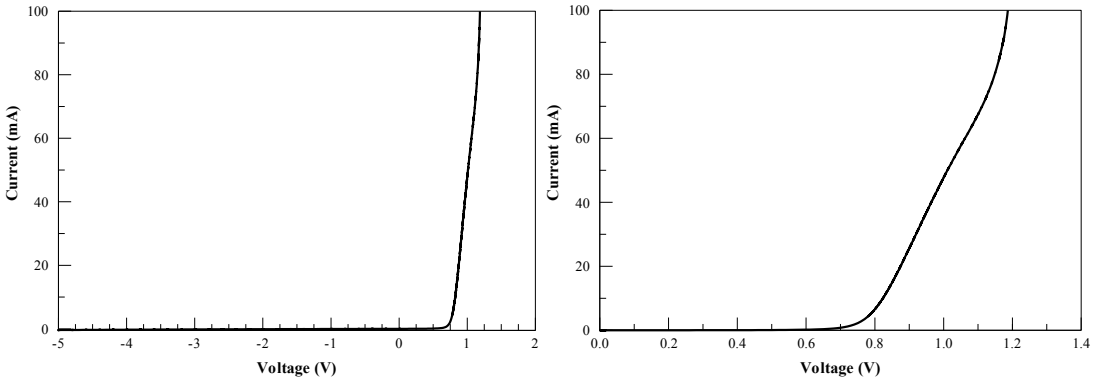


Figure 7.2. Equivalent circuit of the diode (including package parasitics).

C. Measurements

A popular diode used in rectenna design is the M/A-COM GaAs Schottky barrier diode (MA4E1317). This diode will be modeled for use in a 5.8 GHz rectenna system.

The current-voltage curve at DC should be measured so that the DC components of the model (R_s and R_j) can be calculated [94]. This can be done with a curve tracer, or a high resolution voltmeter and ammeter can be used if a curve tracer is not available. The diode must be mounted on a test board to allow the measurement equipment to contact the diode terminals, so care must be taken not to add any extra resistance with the test board. A silver adhesive (Diemat DM6030Hk/F954) is used to connect the diode to the test board. Also, the diode characteristics will be permanently altered if it is burned out. This can happen if over 100 mW is applied to the diode or if the breakdown region is reached. The diode can reach varying degrees of being burned out. Sometimes once burned out, it will become an open or a short, but other times it will just start to permanently behave differently by allowing more current to flow for the same applied voltage. Extra care must be taken to ensure that the diode is not burned out to any degree before the measurements are complete. Seven diodes were measured to ensure that the measurements of a "typical" diode were used to create the model. One of the measured curves of "typical" values can be seen in Fig 7.3.



(a) (b)

Figure 7.3. (a) Diode DC measurements and (b) zoomed in view.

S-parameter measurements of the diode must be taken at numerous bias levels using the Thru-Reflect-Line (TRL) calibration method [95]. A network analyzer is used to measure the S-parameters as ratios of complex voltage amplitudes. These S-parameters should be measured at the diode reference plane as shown in Fig 7.4. However, the measurement plane is at a point within the network analyzer, and losses and phase delays will be introduced to the measurement caused by cables, connections, and transitions that must be in place for the network analyzer to connect to the diode. These losses and phase delays can be lumped into a two-port error box placed at both ports of the diode between the measurement plane and the device plane (Fig 7.5). The TRL calibration method can be used to characterize these error boxes before the measurements of the diode are made. The effects of the error boxes can then be removed from the measurements, leaving only the error-corrected S-parameters of the diode.

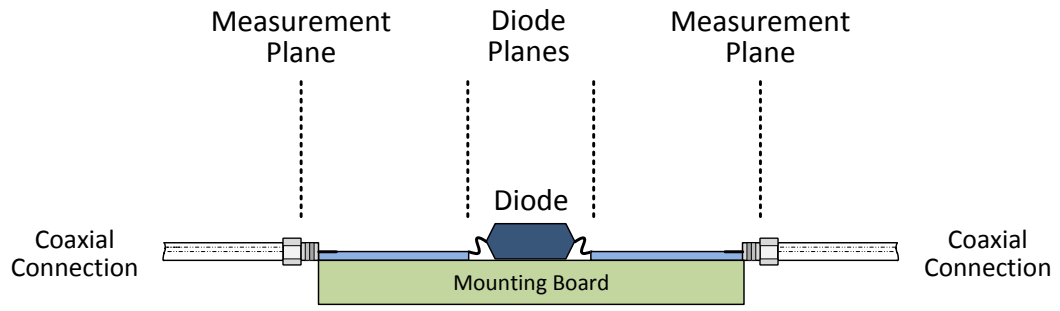


Figure 7.4. Measurement and Device Planes for S-parameter measurements of a diode.

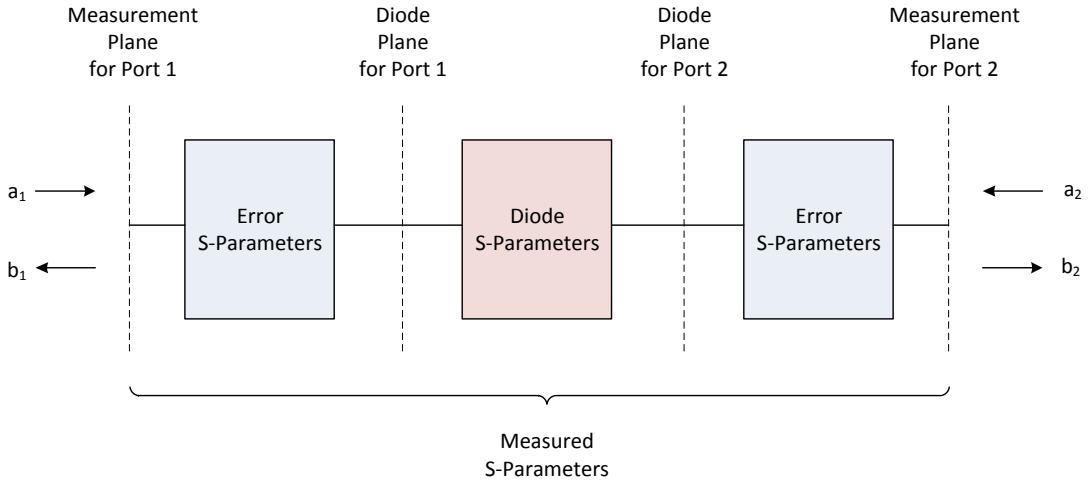


Figure 7.5. Block diagram of a diode's S-parameter measurements.

Three connections are used by the TRL method to determine the error boxes. The Thru connection is made by connecting port 1 directly to port 2 at the device plane. This can be realized by using the same test fixture that will be used for the diode measurement but with the diode removed; the two device planes will be moved infinitesimally close to each other. The Reflect connection is made by terminating each measurement port with a load having a large reflection coefficient. This can be realized by using the same test fixture that will be used for the diode measurement but with the connections to the ports terminated at the device plane. The Line connection is made by connecting port 1 to port 2 through a length of matched transmission line. This can be realized by using the same test fixture as is used for the Thru connection but with an extra length of transmission line added [11]. A diagram of the calibration connections can be seen in Fig 7.6.

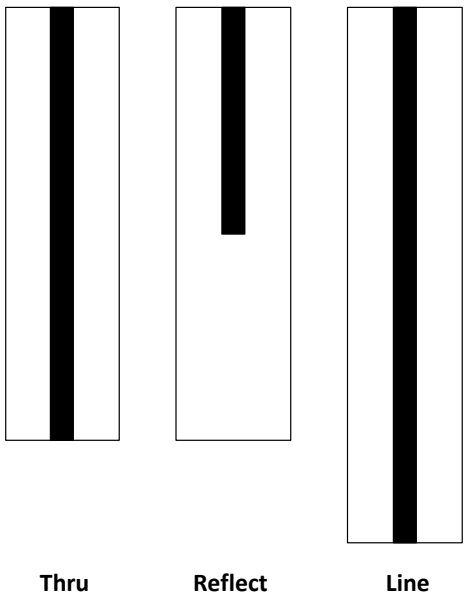


Figure 7.6. Calibration set for TRL measurement.

The S-parameters should be measured over a span of frequencies, but the TRL method places a restriction on the range of frequencies that can accurately be measured. The optimal extra length of the Line (beyond the Thru length) is ¼ wavelength (or 90° of insertion phase) at the center of the frequency range. The insertion phase of the ¼ wavelength Line [96] will vary with frequency as:

$$\text{phase (degrees)} = (360 \times \text{frequency} \times \text{electrical length}) / c \tag{7.4}$$

which can be rearranged to give the electrical length of a ¼ wavelength at the center frequency

$$\text{electrical length (cm)} = 15 / (f_1(\text{GHz}) + f_2(\text{GHz})) \tag{7.5}$$

where f_1 is the start frequency and f_2 is the stop frequency. The insertion phase requirement is that the phase φ over the operating frequency should be $20° \leq \varphi \leq 160°$ [96].

To capture 5.8 GHz and its second harmonic, a frequency range of 4 to 13 GHz is measured which has a center frequency of $f_0 = 8.5$ GHz. The electrical length for this range is calculated from (7.5) as 0.8824 cm. The phase delays from the added ¼ wavelength at the start and stop frequency are calculated from (7.4) as

$$\text{phase (@ 4 GHz)} = 42° \tag{7.6}$$

$$\text{phase (@ 13 GHz)} = 138° \tag{7.7}$$

which meet the insertion phase requirement.

The Thru, Reflect, Line, and numerous diode mounts are etched on 20 mil RT/Duroid 5880 substrate which has a dielectric constant of 2.2. A Thru line length of 110 mm is used, and the physical length of the extra ¼ wavelength on the Line is 6.46

mm. The strip width is 1.52 mm to provide a 50 Ω line at the center frequency. The calibration sets and a diode mount can be seen in Fig 7.7.

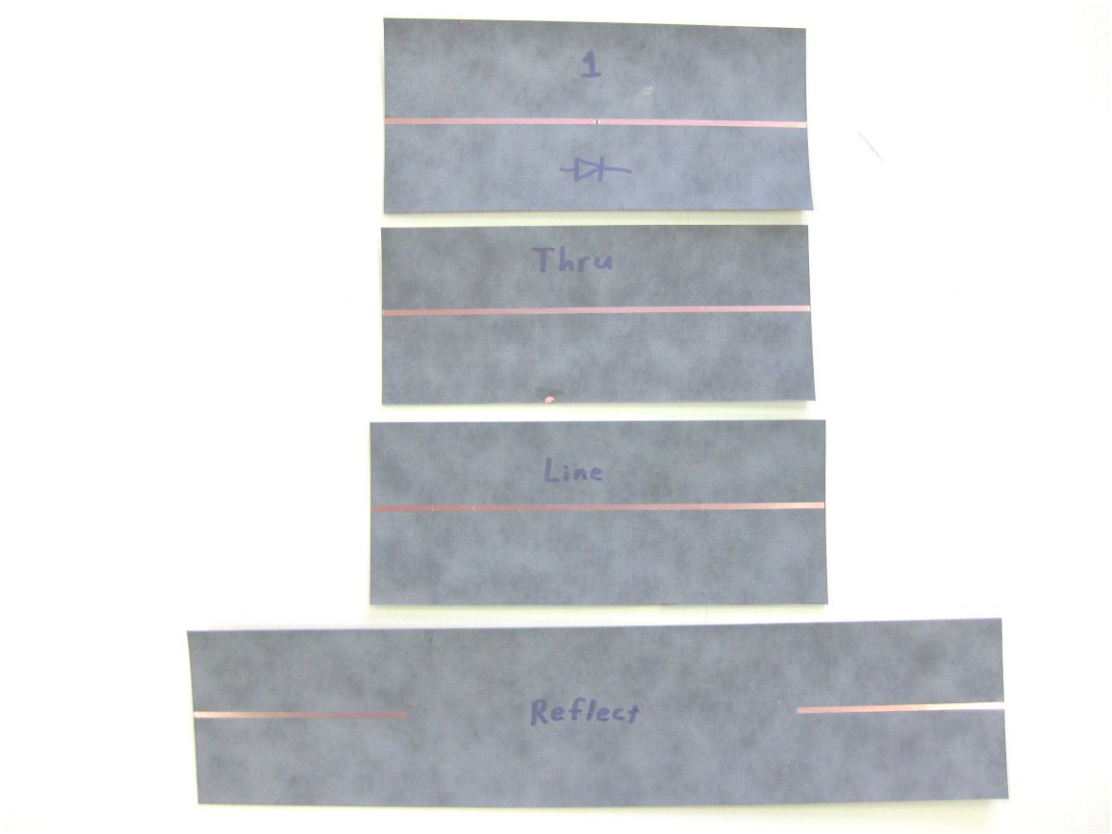


Figure 7.7. Etched calibration set and test fixture for diode TRL measurement.

To calibrate the network analyzer and measure the diode, a measurement setup should be constructed as illustrated in Fig 7.8. Bias-Tees must be used on both sides of the test fixture so that the diode can be biased to various voltages without damaging the network analyzer. The Bias-Tees should be in place (as they will be used for diode measurement) when the calibration sets are used so that the effects of the Bias-Tees are accounted for in the error box calculation. It is important to note that there is some resistance in the Bias-Tees which will result in a voltage drop across the Bias-Tee when a DC current flow is established. Therefore, not all of the voltage provided by the DC-voltage source will appear as a bias across the diode, so a voltmeter should be used to measure the actual bias being applied to the diode.

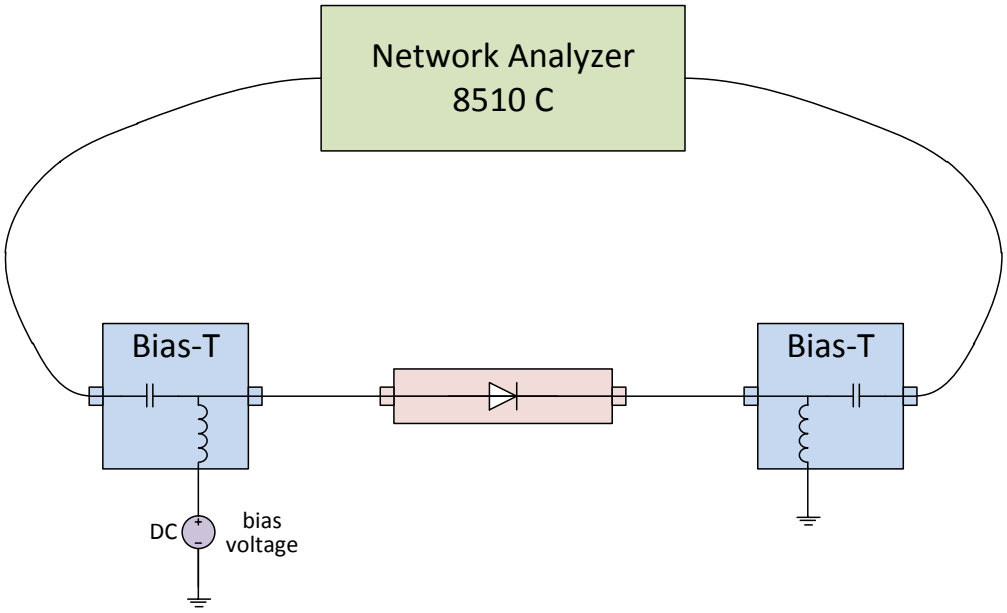
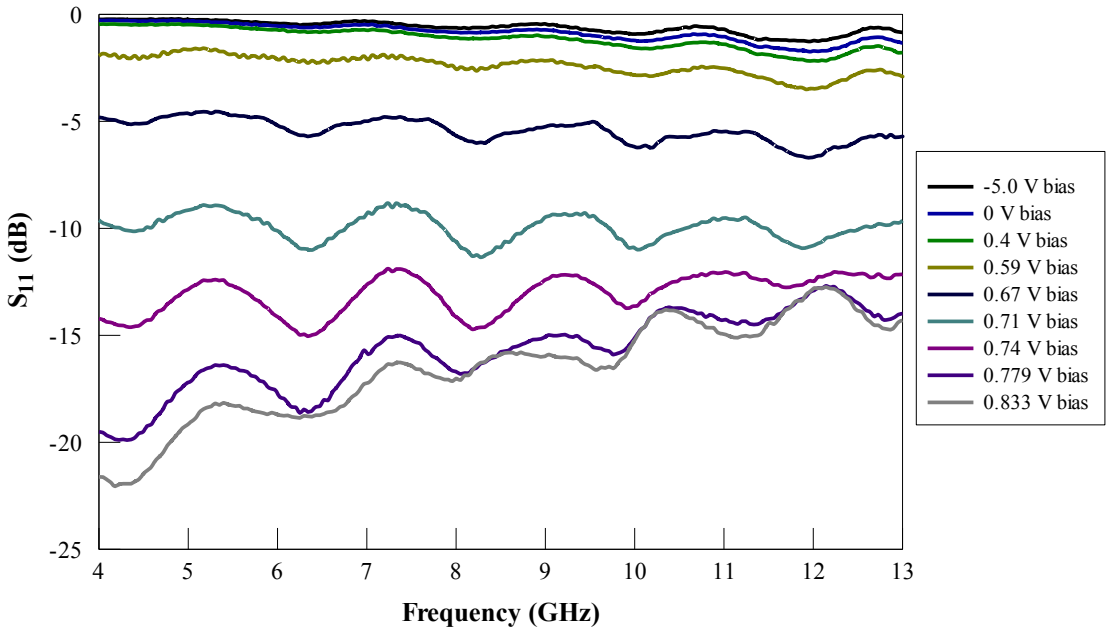
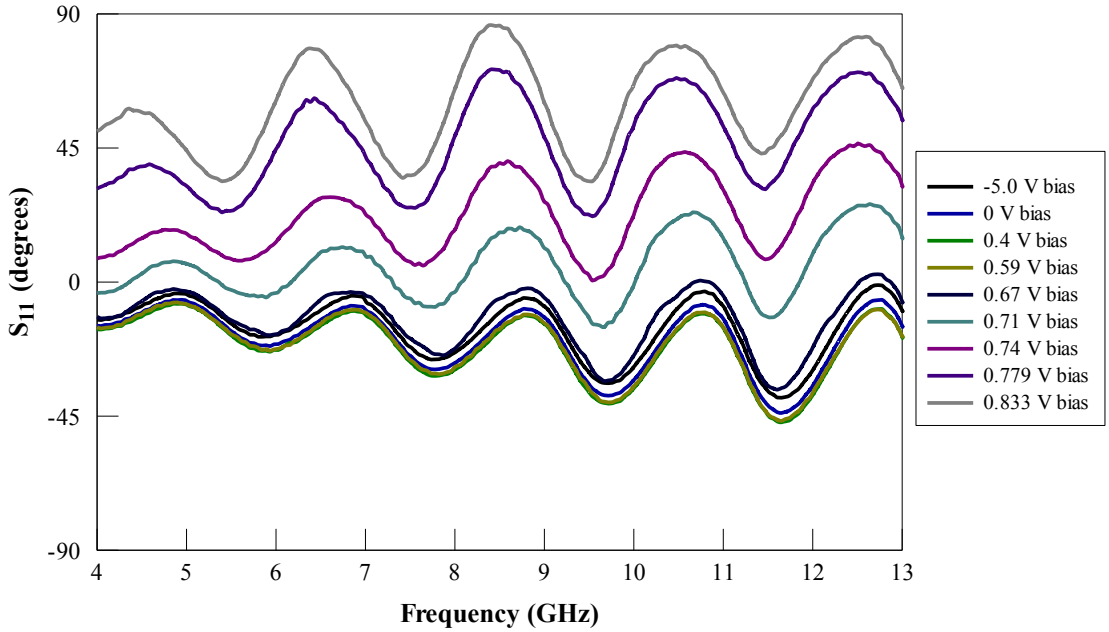


Figure 7.8. Setup for TRL calibrated S-parameter measurement of a diode.

The Agilent 8510C network analyzer can be set to compute the error boxes and directly apply them to the measured S-parameters by entering the correct settings and letting the network analyzer know what kind of calibration set is being used. There is some difficulty in creating consistent connections for all calibration sets and diode mounts. Ten diodes were measured at sixteen bias levels with a Thru length of 110 mm. One of the better measurement sets can be seen in Figs. 7.9-7.10. Some oscillation with frequency can be seen in the figures which is a result of inconsistent connections. Another set of four diodes was measured with a new calibration set which has a Thru length of 37.88 mm in an attempt to reduce some of the transmission line effects on the measurements, but the same oscillation with frequency occurred. The new measurements had fewer oscillations across the frequency range because the shortened length could not support as many wavelengths, but the oscillations had similar amplitude variations. This confirmed that the oscillations were indeed due to connection inconsistencies.

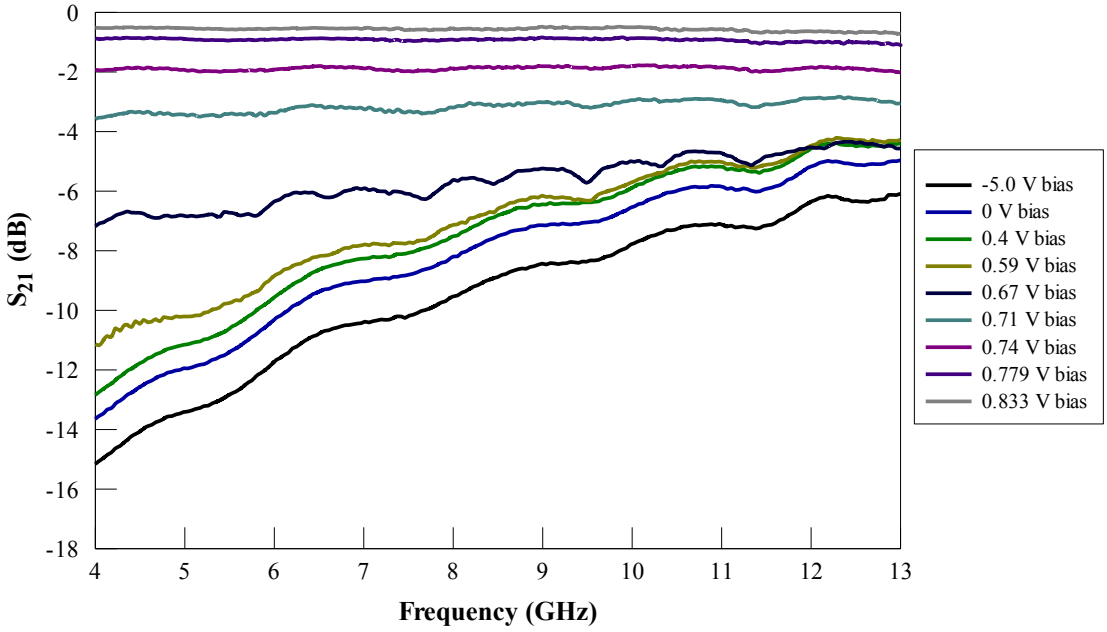


(a)

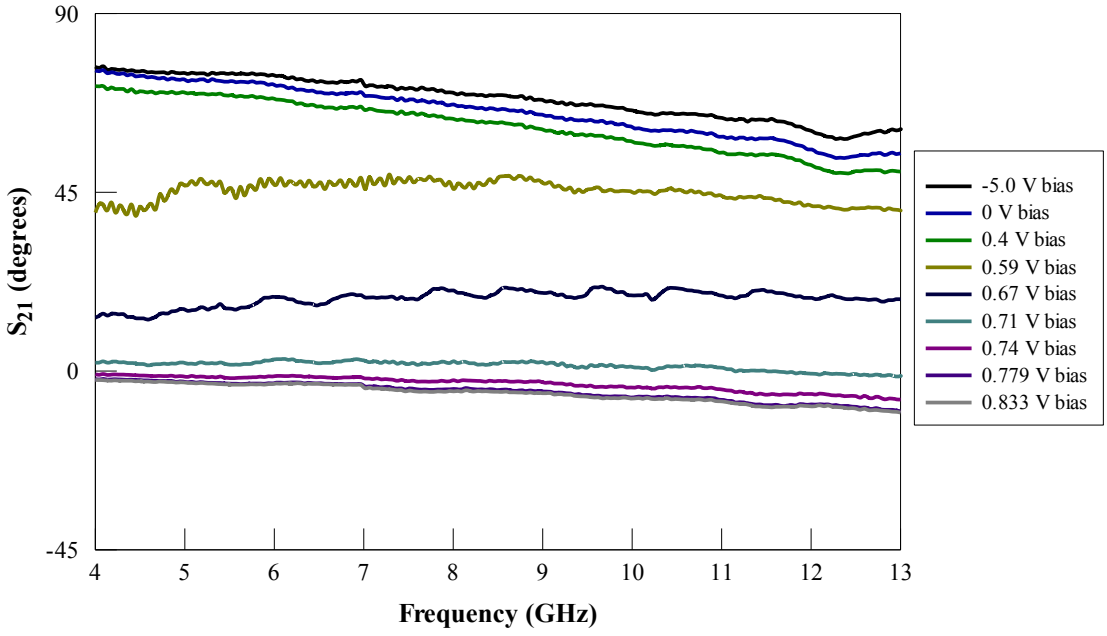


(b)

Figure 7.9. S-parameter (S_{11}) measurements at various bias voltages. (a) Magnitude and (b) phase.



(a)



(b)

Figure 7.10. S-parameter (S_{21}) measurements at various bias voltages. (a) Magnitude and (b) phase.

D. Model Parameter Extraction

Once measurements of the diode's DC current-voltage relation and its small signal S-parameters at various bias levels are successfully taken, the equivalent circuit parameters can be fit to the measurements.

The DC parameters R_s and I_j (interchangeable with R_j) can be fit to the DC current-voltage data because at DC these are the only components with any effect. The equivalent circuit will reduce to Fig 7.11. Note the relation $V_j = V_D - R_s I_j$, where V_D is the measured voltage across the diode and I_j is the measured diode current. The current-voltage curve can be broken down into two regions: the exponential region and the linear region (Fig 7.12) [97]. At low current, the junction resistance R_j is very large. When current increases, the junction resistance decreases, so at some point the junction resistance will become negligibly small. The diode's series resistance will then become the dominant resistor, and the curve will become linear. This means that when current increases, the ohmic effect takes over. In the exponential region, R_j dominates, so

$$I_j = I_0\left(e^{V_j/nV_t} - 1\right) \tag{7.8}$$

can be fit to this region where I_j is used in place of R_j. This curve can be transformed into a linear curve so that linear regression techniques can be utilized:

$$I_j = I_0\left(e^{V_j/nV_t} - 1\right) \tag{7.9}$$

$$\log(I_j) = \log(I_0) + \frac{1}{nV_t}\log(e)\,V_j \ . \tag{7.10}$$

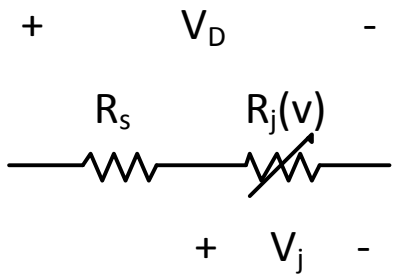


Figure 7.11. Equivalent circuit of the diode at DC.

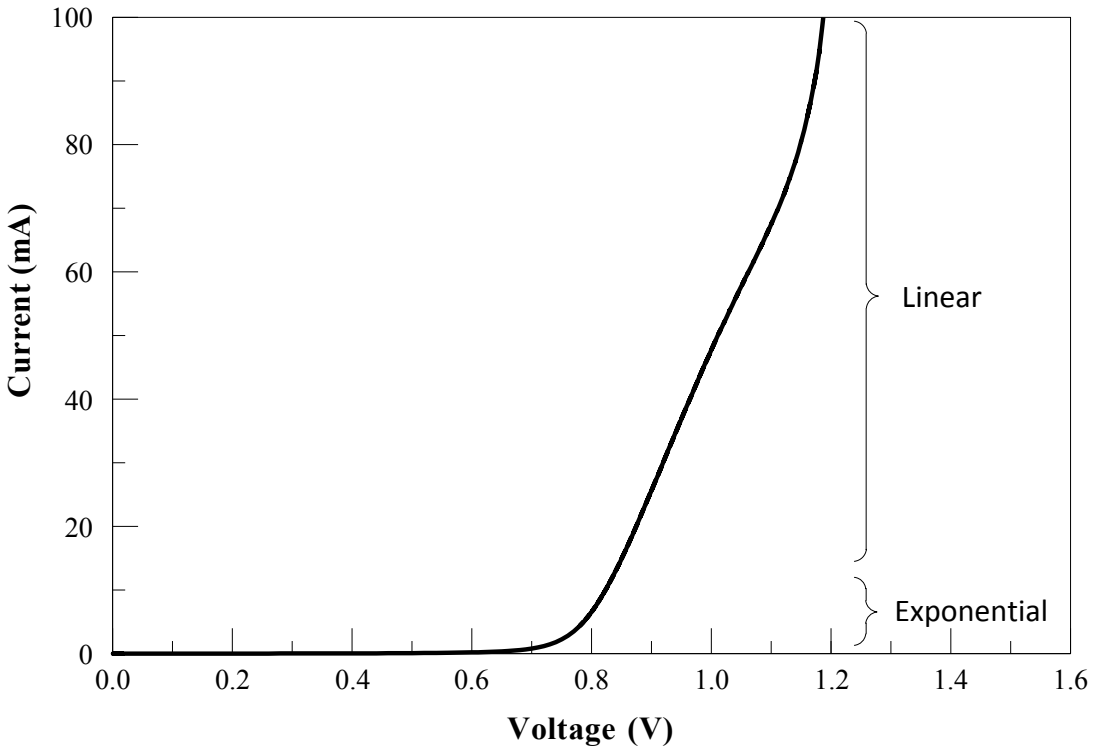


Figure 7.12. Linear and exponential regions of the diode's DC current-voltage curve.

Note that the "-1" in (7.9) is dropped with minimal effect. Equation (7.10) is of the linear form y = m x + b where

$$y = \log(I_j) \tag{7.11}$$

$$b = \log(I_0) \tag{7.12}$$

$$m = \frac{1}{nV_t}\log(e) \tag{7.13}$$

$$x = V_j \tag{7.14}$$

which can be fit to the data in the exponential region using a linear regression. In the linear region, R_s dominates, so

$$R_s = V_D / I_j \tag{7.15}$$

can be fit to this region. R_s is simply the inverse of the curve's slope in the linear region. There is a gradual transition between the exponential region and the linear region, so the slope should be measured at some distance from the exponential region to insure that R_j's effect is minimal. The model parameters are calculated to be:

$$R_s = 4\ \Omega \tag{7.16}$$

$$I_0 = 10^{-14}\ \text{A} \tag{7.17}$$

$$n = 1.06\ . \tag{7.18}$$

The curve is fit with an R^2 value [98] of 0.9901 and is illustrated in Fig 7.13 along with the measured data. Alternatively, a nonlinear regression can be used to fit the curve if capable software is available.

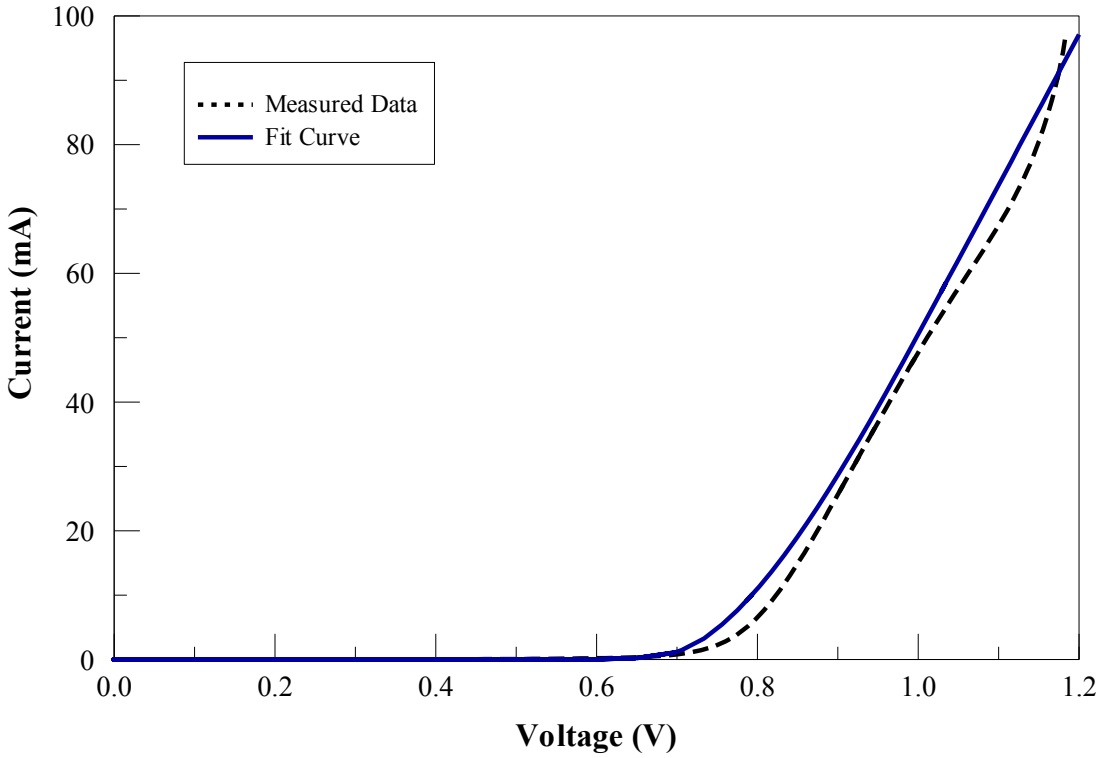


Figure 7.13. Diode DC measurements and fit curve.

Now that R_s and I_j are determined, the other model parameters can be found from the S-parameters. Using Agilent ADS, the S-parameters of the model can be observed for different parameter values. R_s and I_j can be permanently fixed, and then the parasitics can be set to different values while C_j is varied to different bias levels to create a match with measurements over all bias levels. This works successfully when the diode is in the non-conducting region. However, at bias levels in the forward region, the model could not match the measured S-parameters. The restriction of a fixed R_s and I_j prevented the freedom necessary to fully match to the S-parameters.

E. Alternate Model Parameter Extraction Method

It can be difficult to choose correct values for the model parameters, and once parameter values are chosen, they are usually not unique. A convenient method to fit the data using only the S-parameters is to systematically isolate the voltage-independent components from the voltage-dependent components [99]. Circuit components can be added to the diode model to cancel the effect of the voltage-independent components so that only the voltage-dependent components remain as contributing elements. Fig 7.14 illustrates the addition of $-R_s$, $-L_p$, $-C_b$, and $-C_p$ (which are the negatives of the voltage-independent components) to the equivalent circuit which serve to cancel out the voltage-independent components. This leaves only the voltage-dependent components C_j and R_j. When the circuit is simplified to only C_j and R_j in parallel as shown in Fig 7.14, the Y-parameters of this reduced circuit will be a series of straight lines as illustrated in Fig 7.15. The real part of Y_{11} will be $1/R_j(V_j)$ and the imaginary part of Y_{11} will be $\omega C_j(V_j)$ which are both straight lines as a function of frequency.

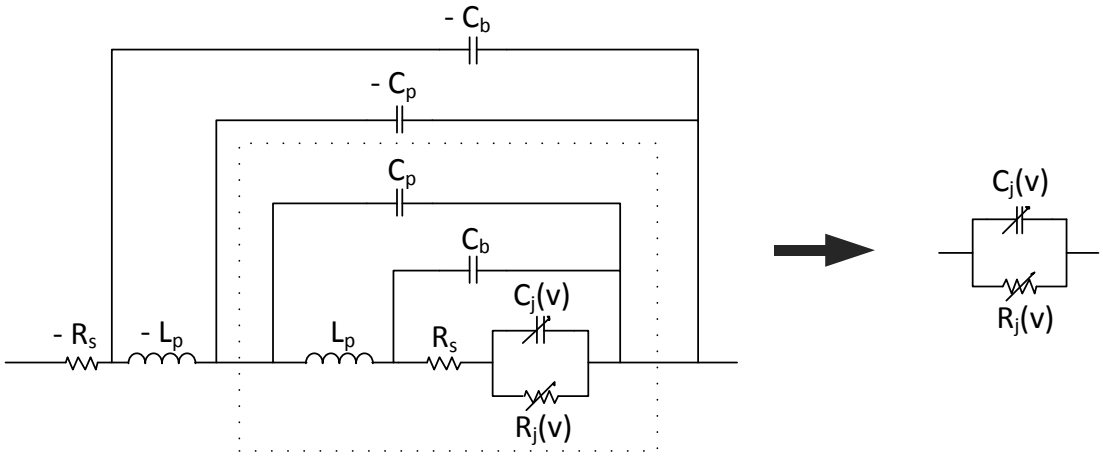


Figure 7.14. Components added to cancel the voltage-independent components' effect. The components within the dotted box represent the diode model. This can be replaced with S-parameters.

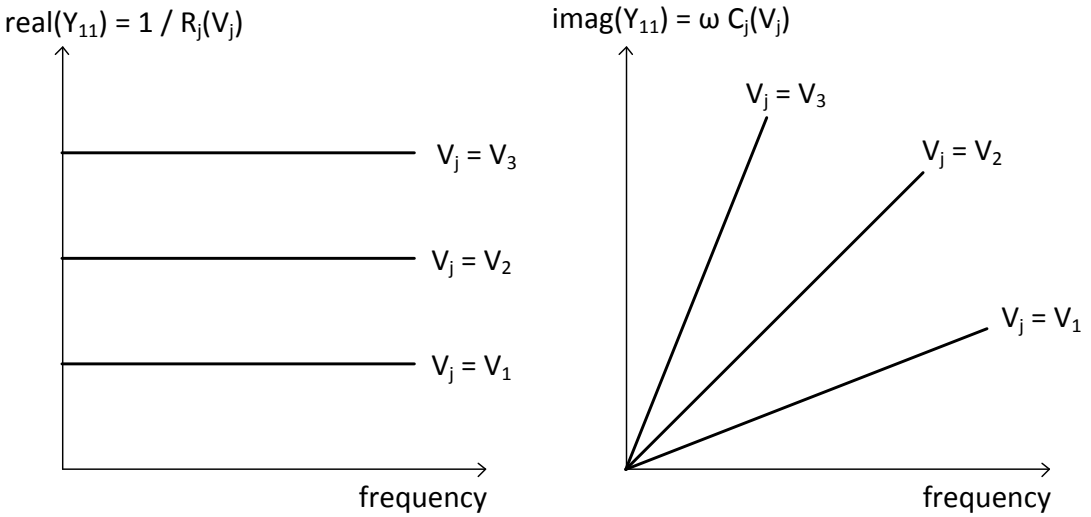


Figure 7.15. Y-parameters of remaining voltage-dependent components. The different lines represent different bias voltages.

Using the Agilent ADS circuit simulator, the equivalent circuit representation of the diode can be replaced by the diode's measured S-parameters and the negative components added as shown in Fig 7.16. The values of Y_{11} can be calculated from the S-parameters [11] by

$$Y_{11} = \left.\frac{I_1}{V_1}\right|_{V_2=0} = Y_0 \frac{(1-S_{11})(1+S_{22})+S_{12}S_{21}}{(1+S_{11})(1+S_{22})-S_{12}S_{21}} \quad . \tag{7.19}$$

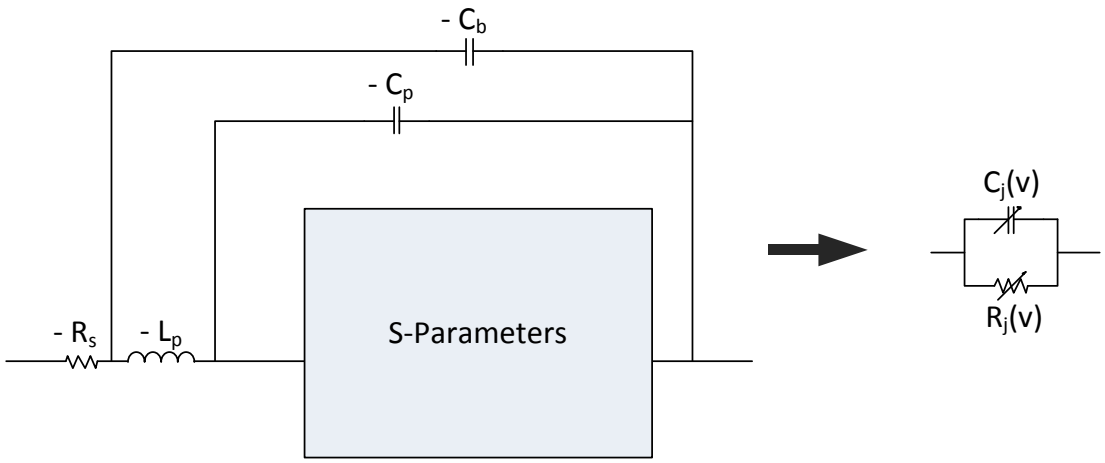


Figure 7.16. Components added to cancel the voltage-independent components' effect. The S-parameters replace the equivalent circuit of the diode. When negative components are set to the correct values, then only C_j and R_j of the diode model will affect the Y-parameters.

If the negative components are set to the correct values, then only C_j and R_j will have any effect on the Y-parameters. Therefore, to find the correct values of the negative components, the values of these components can be varied while observing the resulting Y-parameters. When the Y-parameters are a series of straight lines of the form shown in Fig 7.15, then the effect of the voltage-independent parameters has been cancelled out and their values successfully determined. By following this procedure, these voltage-independent parameters are determined to be:

$$R_s = 0.5\ \Omega \tag{7.20}$$

$$L_p = 0.26\ \text{nH} \tag{7.21}$$

$$C_p = 0.01\ \text{pF} \tag{7.22}$$

$$C_b = 0.04\ \text{pF}\ . \tag{7.23}$$

The values of the voltage-dependent components (as a function of bias voltage) can then be determined from the Y-parameters that result from the previous step. Each straight line on the Y-parameter graph represents a value at a different bias voltage. Since

$$\text{real}(Y_{11}) = 1/R_j(V_j) \tag{7.24}$$

$$\text{imag}(Y_{11}) = \omega C_j(V_j)\ , \tag{7.25}$$

$R_j(V_j)$ can be determined from the real part of Y_{11}, and $C_j(V_j)$ can be determined from the imaginary part of Y_{11}.

F. R_j Calculation

The parameter R_j can be calculated from the real part of Y_{11}. Each horizontal line on the real(Y_{11}) graph gives a value for R_j at a different bias voltage as shown in Fig 7.15. This can be used to create a series of (V_j, R_j) data points (one data point for each bias voltage). The $R_j(V_j)$ data is then converted to a current by

$$I_j = V_j / R_j \,, \tag{7.26}$$

and the equation

$$I_j = I_0 \left(e^{V_j / nV_t} - 1 \right) \tag{7.27}$$

is fit to the data points using a least squares method which gives:

$$I_0 = 1.193 \text{ x } 10^{-11} \text{ A} \tag{7.28}$$

$$n = 1.290 \,. \tag{7.29}$$

The I_j curve is fit with an R^2 value [98] of 0.9812. This curve and the (V_j, I_j) data points are illustrated in Fig 7.17. I_j can be implemented in Agilent ADS as a nonlinear current source by using a symbolically defined device.

A "soft exponential" should be used in place of a regular exponential in the computer calculations to prevent overflow. If the argument of the exponential (V_j/nV_t) is large, taking the exponential of this large argument can cause an overflow error. The "soft exponential" can be defined to be equal to a regular exponential for arguments less than some cutoff value and to be a linear extrapolation of the exponential for arguments greater than the cutoff value. This can be implemented in Agilent ADS as:

max_exp = 1.0e20

max_arg = ln(max_exp)

exp_soft(x) = if (x < max_arg) then exp(x) else (x+1-max_arg)*max_exp endif

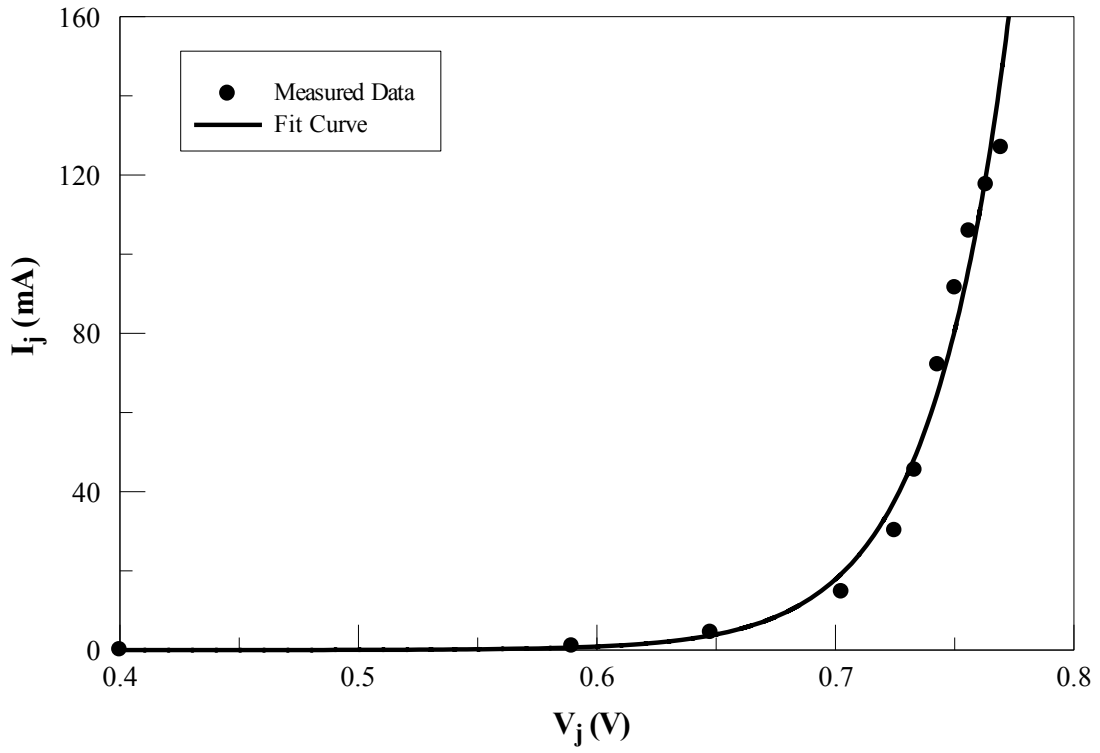


Figure 7.17. I_j curve fit to data.

G. C_j Calculation

The parameter C_j can be calculated from the imaginary part of Y_{11}. Each straight line on the imag(Y_{11}) graph gives a value for C_j at a different bias voltage as shown in Fig 7.15. This can be used to create a series of (V_j, C_j) data points (one data point for each bias voltage). Agilent ADS has a nonlinear capacitor device which can be used to implement C_j. This device requires input in the form of a polynomial, so the $C_j(V_j)$ data must be forced into the form of a polynomial [100].

A polynomial of a large degree (20-25) must be used in order to satisfactorily force the polynomial to behave as an exponential over the desired range, but only sixteen data points exist. If a 20-25 degree polynomial is fit to only sixteen data points using a least squares method then the data is overfit, and the polynomial will fit correctly at the data points but will wildly oscillate between data points. This does not adequately represent the exponential form of the data. Artificial data points can be created by linearly interpolating between real data points, allowing the use of as many data points as desired. A 20-25 degree polynomial can then be fit to the artificial data points (using a least squares method) and effectively fit the form of an exponential within the desired range. Note that the polynomial blows up outside of the range. Polynomials of such a large degree are required to adequately fit the form of an exponential since this is not a polynomial's usual form. A 20 degree polynomial and a 25 degree polynomial are fit to the data using this method with resulting R^2 values [98] of 0.7152 and 0.9339, respectively. These curves and the (V_j, C_j) data points are illustrated in Fig 7.18.

The polynomials are:

20 deg $C_j(V_j)$ =

$2.251953845934267 \times 10^{-14} + 3.928708106450134 \times 10^{-14} V_j - 1.872605744144919 \times 10^{-13} V_j^2 - 8.010052760857773 \times 10^{-13} V_j^3 + 1.035723793272987 \times 10^{-12} V_j^4 + 6.019608944715903 \times 10^{-12} V_j^5 + 3.292217192099333 \times 10^{-12} V_j^6 - 1.188012141308037 \times 10^{-11} V_j^7 - 2.021997434144966 \times 10^{-11} V_j^8 - 6.795781565984654 \times 10^{-12} V_j^9 + 1.408246385109605 \times 10^{-11} V_j^{10} + 2.231293123305528 \times 10^{-11} V_j^{11} + 1.691875529472090 \times 10^{-11} V_j^{12} + 8.293398780762461 \times 10^{-12} V_j^{13} + 2.844367116709308 \times 10^{-12} V_j^{14} + 7.004768751099705 \times 10^{-13} V_j^{15} + 1.238874503528577 \times 10^{-13} V_j^{16} + 1.540283747327346 \times 10^{-14} V_j^{17} + 1.280357467024890 \times 10^{-15} V_j^{18} + 6.396262419195184 \times 10^{-17} V_j^{19} + 1.453338557125182 \times 10^{-18} V_j^{20}$.

25 deg $C_j(V_j)$ =

$1.646738462080980 \times 10^{-14} - 3.841968572067818 \times 10^{-14} V_j + 1.353667676256704 \times 10^{-13} V_j^2 + 1.603431782387508 \times 10^{-12} V_j^3 + 1.725111435338492 \times 10^{-13} V_j^4 - 1.363440404577148 \times 10^{-11} V_j^5 - 1.692864176044302 \times 10^{-11} V_j^6 + 3.576722607831512 \times 10^{-11} V_j^7 + 9.264821210126666 \times 10^{-11} V_j^8 + 2.660693196199393 \times 10^{-11} V_j^9 - 1.334940731727817 \times 10^{-10} V_j^{10} - 1.929241277273865 \times 10^{-10} V_j^{11} - 7.342053211680040 \times 10^{-11} V_j^{12} + 9.197717732432363 \times 10^{-11} V_j^{13} + 1.623743979314330 \times 10^{-10} V_j^{14} + 1.323625389434068 \times 10^{-10} V_j^{15} + 7.171379278258487 \times 10^{-11} V_j^{16} + 2.826579043343990 \times 10^{-11} V_j^{17} + 8.387264303460990 \times 10^{-12} V_j^{18} + 1.895653489292848 \times 10^{-12} V_j^{19} + 3.256194634903012 \times 10^{-13} V_j^{20} + 4.188460909887590 \times 10^{-14} V_j^{21} + 3.913242592828937 \times 10^{-15} V_j^{22} + 2.510211233005998 \times 10^{-16} V_j^{23} + 9.894061095862421 \times 10^{-18} V_j^{24} + 1.807503156397230 \times 10^{-19} V_j^{25}$.

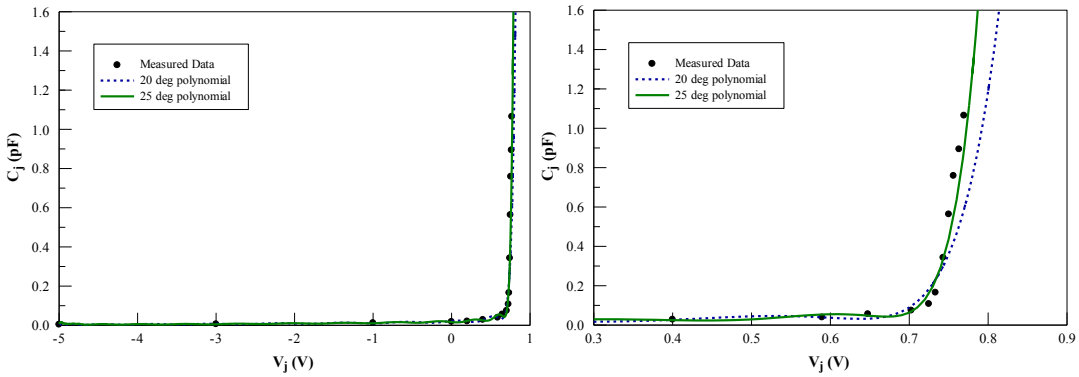


Figure 7.18. (a) C_j polynomials (of 20 and 25 degree) fit to data and (b) zoomed in view.

The equivalent circuit can now be constructed in Agilent ADS by using a current source to represent R_j and a nonlinear capacitor to represent C_j. This equivalent circuit shows a good match to the measured S-parameters with a slight S-parameter oscillation with changing bias voltage; this is due to the oscillatory properties of the polynomial which was used to represent C_j. This might be acceptable, however a major problem results when a harmonic balance simulation is run using this model. There is a huge convergence problem which is a byproduct of the nonlinear capacitor [101]-[102]. This makes the model very difficult to use for the design of a rectenna and severely limits its utility as a design tool.

H. C_j Calculation (Charge-Based Model)

A charge-based model can be used to represent the voltage-dependent capacitor in place of the nonlinear capacitor component to correct the convergence problem in the harmonic balance simulation [103]. The equation for junction capacitance in the reverse bias region has been previously defined as

$$C_{jr}(V_j) = C_0 \sqrt{\frac{V_0}{V_0 - V_j}} \tag{7.30}$$

where C_{jr} signifies that this is valid only for the reverse bias region. Since

$$Q(v) = \int_v C(\hat{v})\, d\hat{v} + Q_0 \ , \tag{7.31}$$

(7.30) can be integrated with respect to V_j to obtain an equation for charge:

$$Q_{jr}(V_j) = -2C_0 \sqrt{V_0 (V_0 - V_j)} \tag{7.32}$$

where a boundary condition of $Q_j(V_0) = 0$ is enforced. $C_{jr}(V_j)$ and $Q_{jr}(V_j)$ can be extended into the forward region using linear extrapolation. A variable α (where $0<\alpha<1$) must be introduced to define the reverse region as $V_j \leq \alpha V_0$ and the forward region as $V_j > \alpha V_0$. This is necessary to extend the range of the equations into the forward-region while preventing $C_{jr}(V_j)$ from blowing up at $V_j = V_0$. $C_{jr}(V_j)$ is extended to the forward-region by linear extrapolation as

$$C_{jf}(V_j) = C_{jr}(\alpha V_0) + C_{jr}'(\alpha V_0)(V_j - \alpha V_0) \ , \tag{7.33}$$

which can be integrated with respect to V_j to obtain an equation for charge in the forward-region:

$$Q_{jf}(V_j) = \left[V_j - \alpha V_0\right]\left[C_{jr}(\alpha V_o)\right] + C'_{jr}(\alpha V_0)\left[\frac{(V_j - \alpha V_0)^2}{2}\right] + Q_{jr}(\alpha V_0) \qquad (7.34)$$

where a boundary condition of $Q_{jf}(\alpha V_0) = Q_{jr}(\alpha V_0)$ is enforced. This can be rewritten as

$$Q_{jf}(V_j) = \frac{C_0}{\sqrt{1-\alpha}}\left(V_j - \alpha V_0 + \frac{(V_j - \alpha V_0)^2}{4V_0(1-\alpha)}\right) + Q_{jr}(\alpha V_0)\ . \qquad (7.35)$$

So the final equations for $C_j(V_j)$ and $Q_j(V_j)$ are defined as

$$C_j(V_j) = \begin{cases} C_{jr}(V_j)\ ,\ V_j \le \alpha V_0 & \text{(reverse bias)} \\ C_{jf}(V_j)\ ,\ V_j > \alpha V_0 & \text{(forward bias)} \end{cases} \qquad (7.36)$$

$$Q_j(V_j) = \begin{cases} Q_{jr}(V_j)\ ,\ V_j \le \alpha V_0 & \text{(reverse bias)} \\ Q_{jf}(V_j)\ ,\ V_j > \alpha V_0 & \text{(forward bias)} \end{cases} \qquad (7.37)$$

with $C_{jr}(V_j)$, $C_{jf}(V_j)$, $Q_{jr}(V_j)$, and $Q_{jf}(V_j)$ defined in (7.30), (7.33), (7.32), and (7.35) respectively. It is important to note that $C_j(V_j)$ and $Q_j(V_j)$, as well as their first derivatives, are continuous. $C_j(V_j)$ can be fit to the data points using a least squares method which gives:

$$C_0 = 1.771 \text{ x } 10^{-14} \text{ F} \qquad (7.38)$$

$$V_0 = 0.7424 \text{ V} \qquad (7.39)$$

$$\alpha = 0.9942\ . \qquad (7.40)$$

The C_j curve is fit with an R^2 value [98] of 0.9985. Note that in addition to correcting the convergence problem, the charge-based model allows a more natural curve to be fit and increases C_j's match to the data. The fit curve for C_j and the (V_j, C_j) data points can be seen in Fig 7.19. C_j can be implemented in Agilent ADS with a symbolically defined device. A weighting function of 1 in an ADS symbolically defined

device is defined as jω and is used when a time derivative is required. Q_j is defined in (7.37) as a function of C_0, V_0, and α, and since

$$I = \frac{d}{dt}\big(Q(v)\big) \ , \tag{7.41}$$

a weighting function of 1 can be used on an expression for Q_j to define the current through the nonlinear capacitor. This is the charge-based model of a voltage-dependent capacitor.

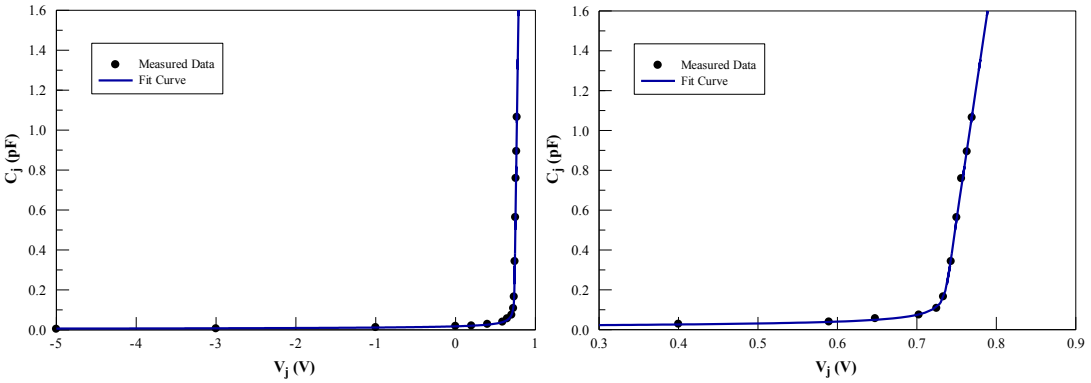


Figure 7.19. (a) C_j exponential curve fit to data and (b) zoomed in view.

I. Final Model

The final diode model is created using equations (7.27) and (7.37) with parameter values (7.20)-(7.23), (7.28)-(7.29), and (7.38)-(7.40). The implementation of this model in Agilent ADS [83] can be seen in Fig 7.20. Note that the variable capacitor C_j and variable resistor R_j are both implemented as symbolically defined devices.

The S-parameters of the model show a good match to the measured S-parameters; a comparison can be seen at a few of the bias voltages in Figs. 7.21-7.22. Sixteen bias levels were measured, but only a few are shown in order to make the graphs more readable. This diode model can be used in an electromagnetic simulator, such as Agilent ADS, in conjunction with the other rectenna components (which are easily modeled) to simulate a rectenna's performance. This will greatly improve rectenna design time and prevent unnecessary fabrication and testing. An example of using this model for the design of a rectenna system is included in the next section.

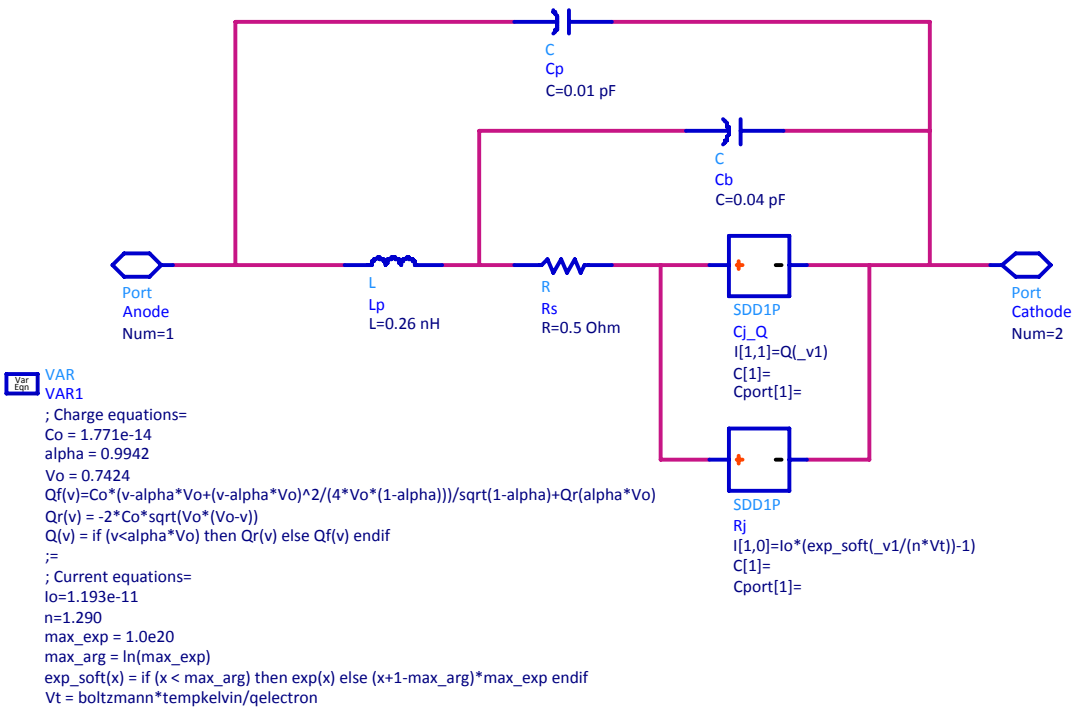
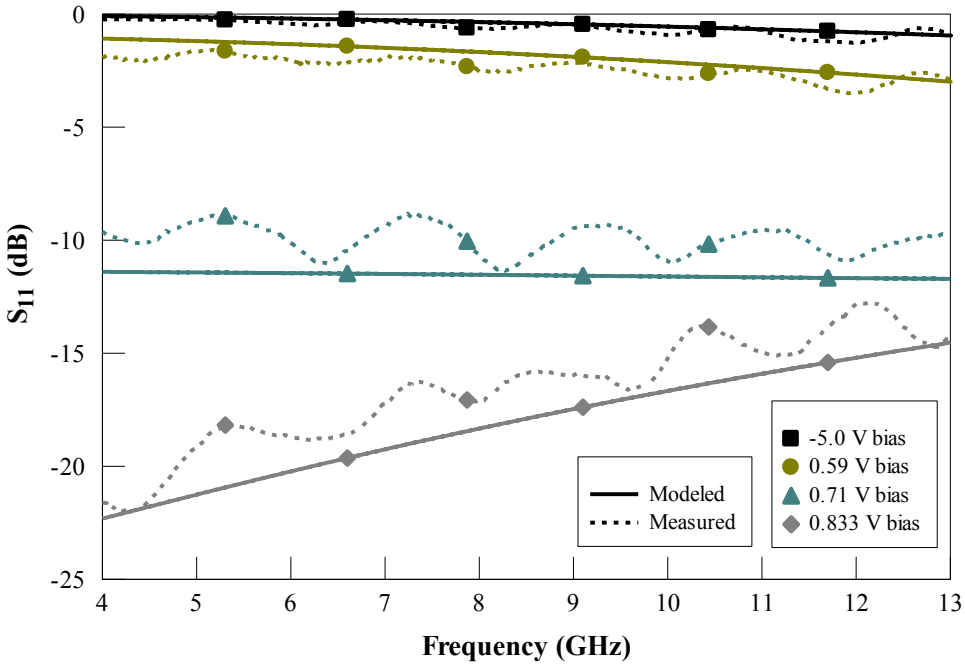
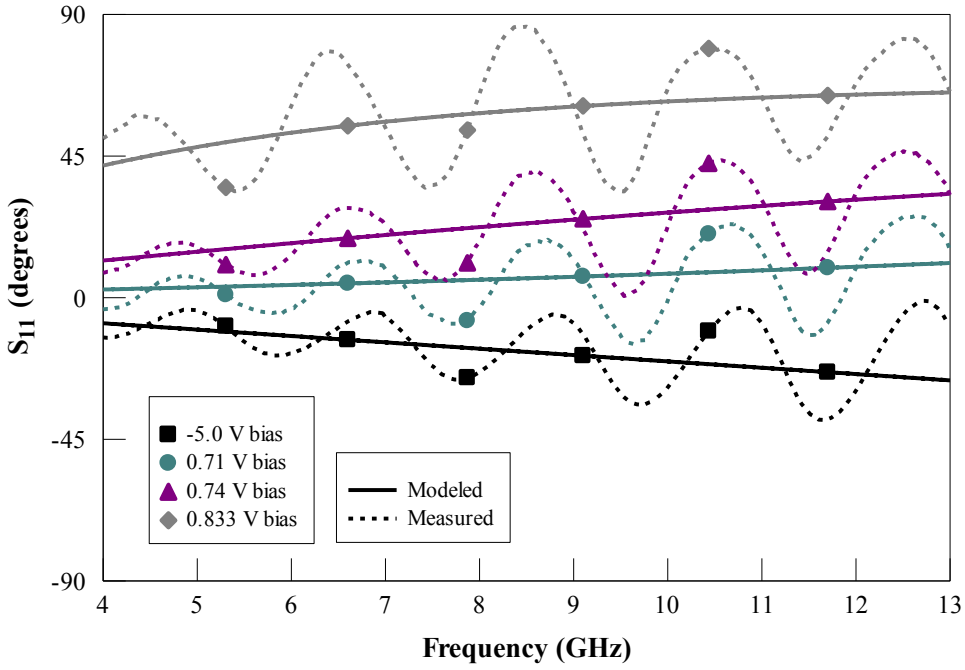


Figure 7.20. Schematic of the diode model in Agilent ADS.

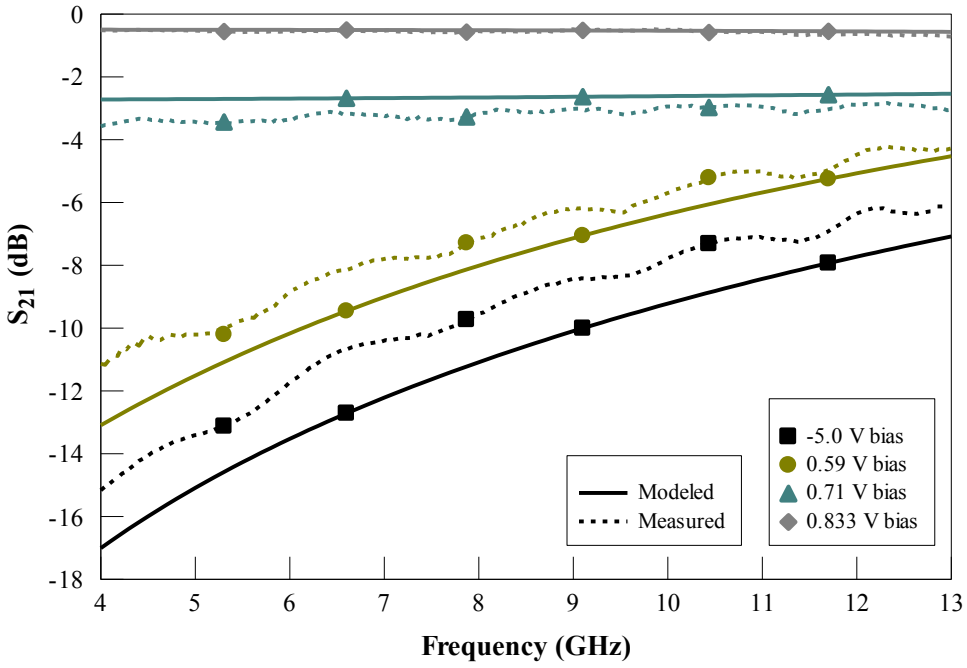


(a)

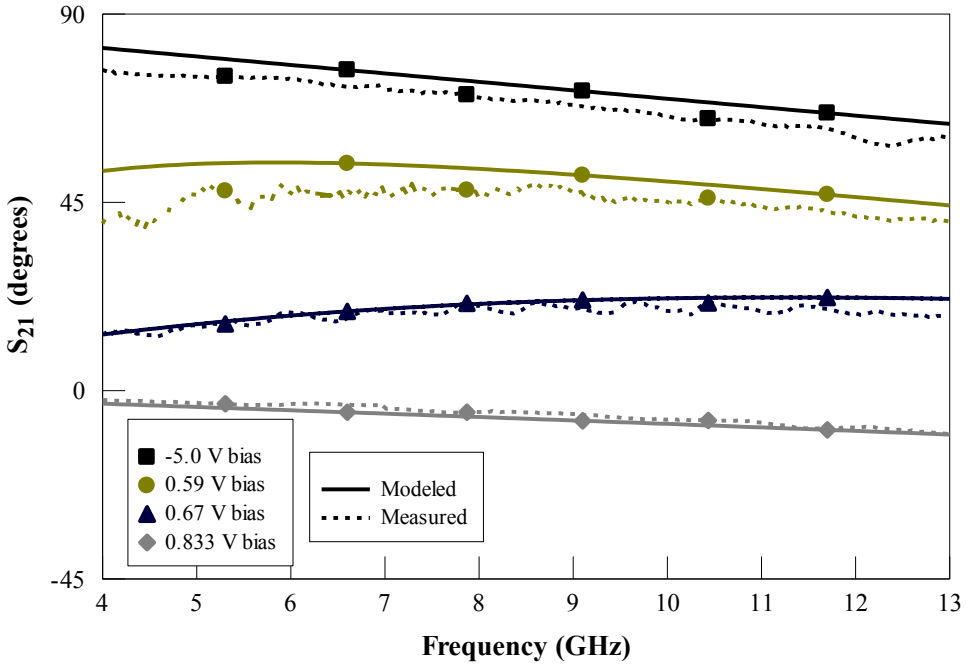


(b)

Figure 7.21. Comparison of S-parameter (S_{11}) measurements and modeled (simulated) values at various bias voltages. (a) Magnitude and (b) phase.



(a)



(b)

Figure 7.22. Comparison of S-parameter (S_{21}) measurements and modeled (simulated) values at various bias voltages. (a) Magnitude and (b) phase.

J. Conclusion

A diode model based on an equivalent circuit was created to aid in rectenna design. A method which systematically isolates the model's voltage-dependent components from its voltage-independent components was used to calculate the parameters. This method is based on measured S-parameters of the diode at various bias levels. The voltage-independent components were found by adjusting their values such that the Y-parameters appeared as a series of straight lines. The voltage-dependent components were then derived by fitting curves to data points obtained from the straight-lined Y-parameters. A charge-based capacitor model was used in lieu of an ADS nonlinear capacitor component to avoid harmonic balance convergence issues.

The model was implemented in Agilent ADS using symbolically defined devices for both the voltage-dependent current source and the voltage-dependent capacitor (implemented as a charge source). The simulated S-parameters of the model showed a good match to the measured S-parameters over all the bias levels. This model can be used in ADS to simulate the entire rectenna design as will be shown in the next section.

8. RECTENNA DESIGN *

A. Design and Fabrication

The most important component of a microwave wireless power transmission (WPT) system is the rectenna which consists of a receiving antenna and a rectifying circuit. The antenna receives the microwave power, and the rectifying circuit converts the microwave power into DC power.

The proposed rectenna design for 5.8 GHz operation is illustrated in Fig. 8.1. It consists of an aperture-coupled high-gain single-patch antenna, a split ring resonator (SRR) filter to suppress harmonic signals, a diode for RF-to-DC conversion, a DC-pass filter (capacitor), and a load.

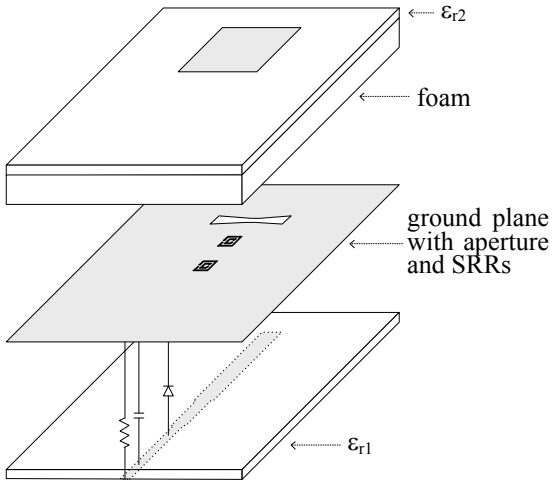


Figure 8.1. Rectenna design (exploded view).

* Parts of this section are reprinted, with permission, from J. Hansen, C. Ahn, and K. Chang, "A 5.8 GHz high gain, aperture coupled rectenna utilizing a split ring resonator filter," in *IEEE Antennas and Propagation Society International Symposium (APSURSI)*, pp. 1-4, Toronto, July, 2010. Copyright © 2010, IEEE.

The role of the antenna is to efficiently receive microwave power from free space. A high gain antenna is important because it will allow the rectenna to receive more power for a given incident power due to its increased effective aperture area. This will allow for longer distance WPT and a smaller physical receiving area. A patch is utilized because of its simplicity, flatness, and ease of fabrication and integration [1][104]. The length of the patch can be initially approximated [5] by

$$L \approx 0.49\lambda_d = 0.49\frac{\lambda_0}{\sqrt{\varepsilon_r}} \tag{8.1}$$

and optimized using IE3D [12]. The antenna must also be tuned to maximize its gain. The antenna is measured to have a gain of 9.3 dBi, and the radiation pattern can be seen in Fig. 8.2.

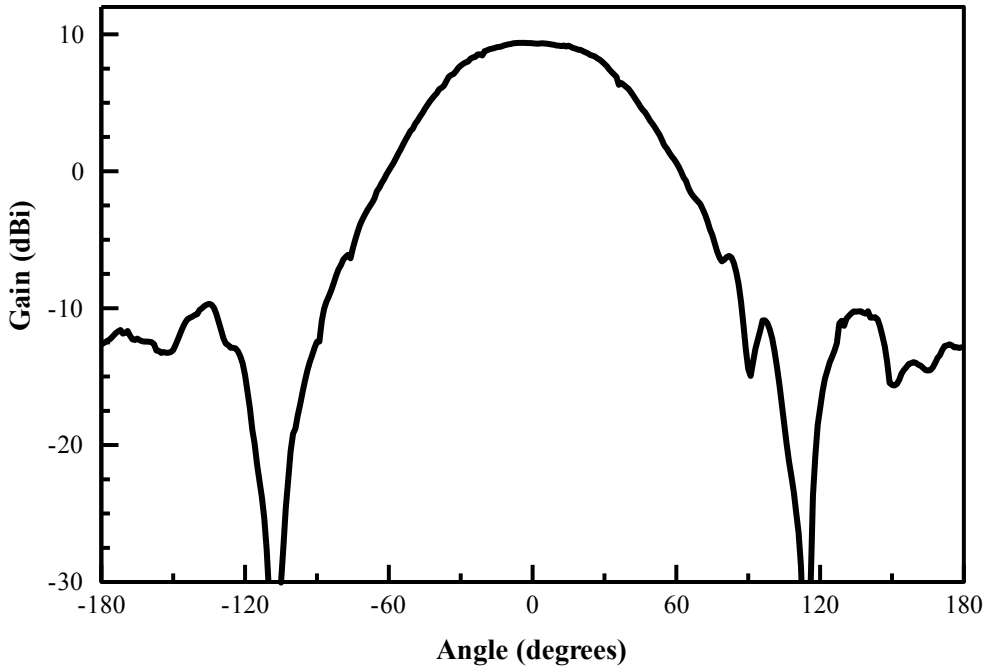


Figure 8.2. Measured radiation pattern.

An aperture-coupled feed [105] is employed because of its inherent isolation between the feeding circuit and the patch. This serves to isolate the antenna operation from the rectifying operation and prevents any inadvertent coupling between the two [1][104]. This will also serve to simplify the design process since the antenna and the rectifier can be designed independently due to very limited coupling between the two.

Novel compact split ring resonators (SRRs) are utilized as the band-reject filter to pass 5.8 GHz and block the second harmonic at 11.6 GHz. This is necessary since the diode creates harmonics due to its non-linear behavior. Most of this harmonic power is contained in the second harmonic, so blocking this harmonic is the most important. The return loss of the patch and filter combination is measured to be 0.86 dB at the second harmonic while the return loss of the patch alone is 5.79 dB at the second harmonic (Fig. 8.3). This is an additional 4.93 dB of suppression provided by the filter.

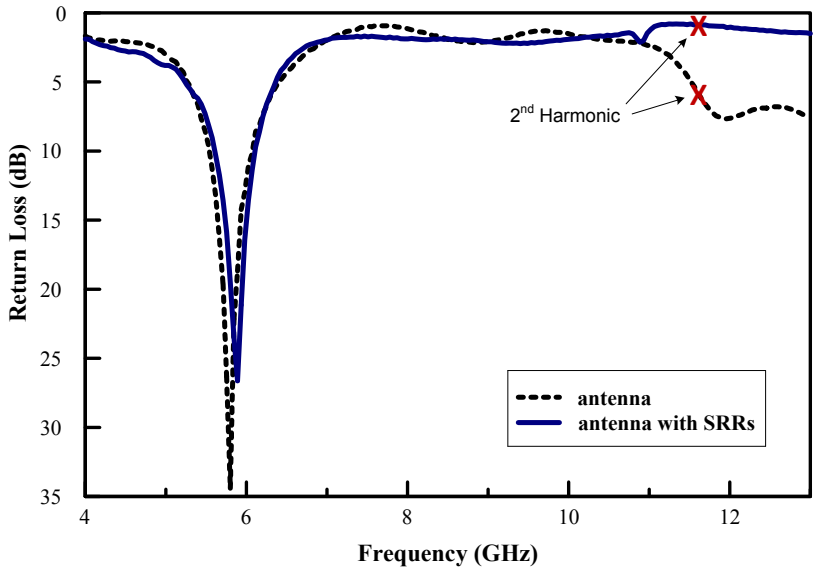


Figure 8.3. Measured return loss of the antenna with and without the split ring resonator (SRR) filter.

The split ring resonator is a defective ground structure which is implemented by etching the structure into the ground plane. This disturbs the shield current in the ground plane and changes the characteristics of the transmission line. A magnetic resonance is induced in the rings of the SRR by the split in the ring and by the gap between the two rings. The splits in the rings allow this type of structure to support wavelengths much larger than the diameter of the rings; this makes SRRs a compact filter [106]-[107]. A split ring resonator is illustrated in Fig. 8.4. The dimensions used for this filter are: a = 1.89 mm, b = 3.01 mm, g = 0.19 mm, t = 0.24 mm, and c = 0.15 mm.

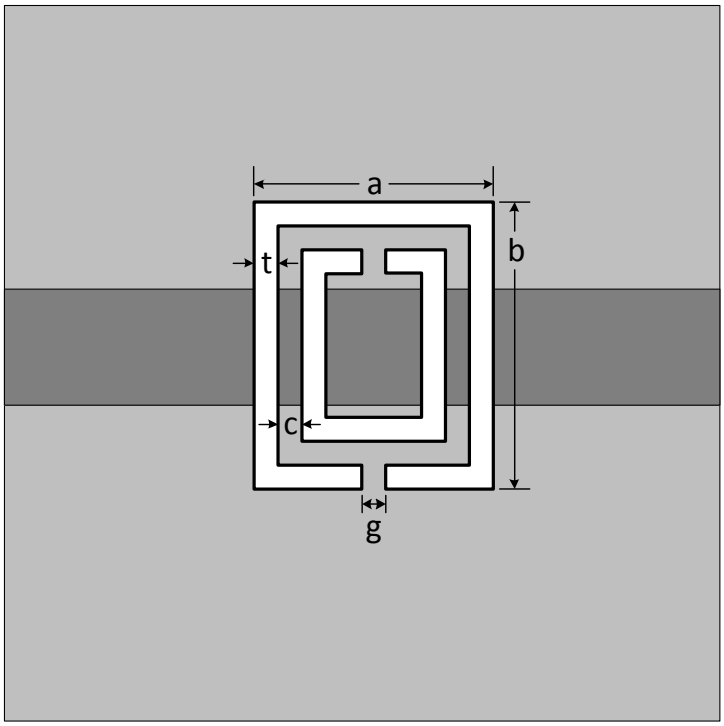


Figure 8.4. Split ring resonator (SRR) in the defective ground plane. The transmission line is also illustrated below the SRR.

A flip-chip type GaAs Schottky barrier diode (MA4E1317) is used to rectify the microwave power, and a broadband capacitor (C08BLBB1X5UX) is used as a DC-pass filter to reflect any unconverted microwave power back to the diode and prevent microwave power from reaching the load. A silver adhesive (Diemat DM6030Hk/F954) is used to attach both the diode and capacitor to the microstrip line. The distance between the diode and the capacitor is crucial because in addition to reflecting the microwave power, the capacitor is used to tune out the diode's reactance and improve conversion efficiency. Setting the distance between the diode and capacitor to a quarter-wavelength will make the capacitor appear as either an open or a short at RF for the fundamental frequency and all of the harmonics. This will create a situation where the microwave energy is reflected back from the capacitor. The SRRs will also reflect the harmonics on the other side of the diode, so any harmonics should be reflected between the SRRs and the capacitor until they are eventually converted to DC by the diode.

Optimization of the rectifying circuit is important to maximize efficiency. The diode must be matched to the input power level and to the load. It has been found that this diode matches well with a 120 Ω line, so the 50 Ω line is converted to a 120 Ω line by way of a quarter-wavelength transformer after the SRRs. The characteristic impedance of the quarter-wavelength transformer can be found [3] by

$$Z_{0T} = \sqrt{Z_i Z_t} \tag{8.2}$$

where Z_{0T} is the characteristic impedance of the quarter-wavelength transformer, Z_i is the input impedance (50 Ω), and Z_t is the terminating impedance (120 Ω). The optimum load value should also be selected to maximize efficiency.

The rectenna is designed using IE3D [12] and is illustrated in Fig. 8.5. The feedline and rectifying circuit are etched on one side of a 20 mil RT/Duroid 5880 substrate with a dielectric constant of 2.2, and the aperture is etched on the other side. Via holes are necessary to connect the diode and capacitor to ground. The patch element is etched on a 2 mil RT/Duroid 3850 substrate with a dielectric constant of 2.9 and is supported by a 3.2 mm thick layer of lightweight foam (Cuming C-FOAM PF-4/PSA) with a low (1.06) dielectric constant. Photographs of the rectenna can be seen in Fig. 8.6.

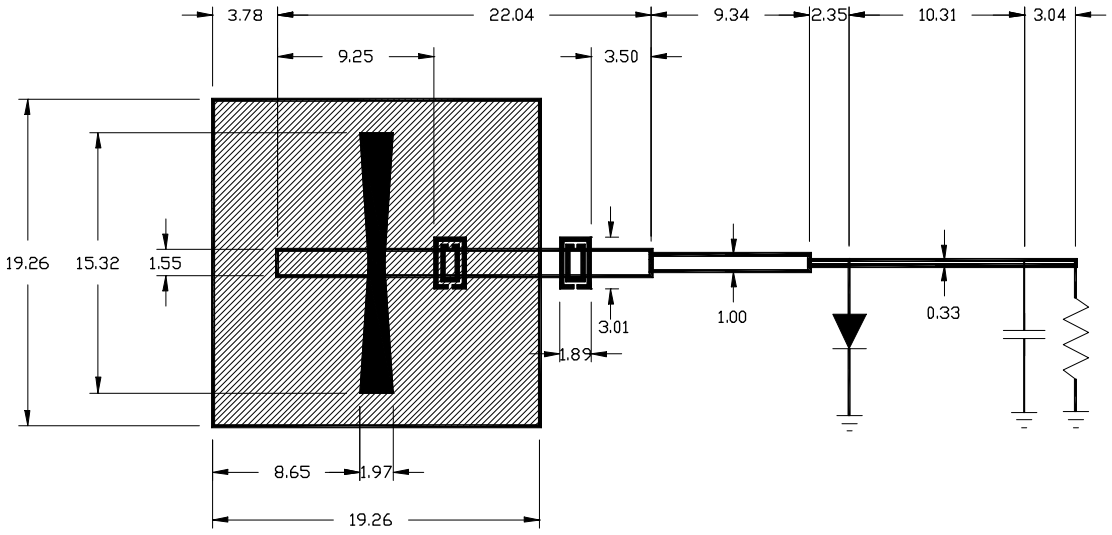
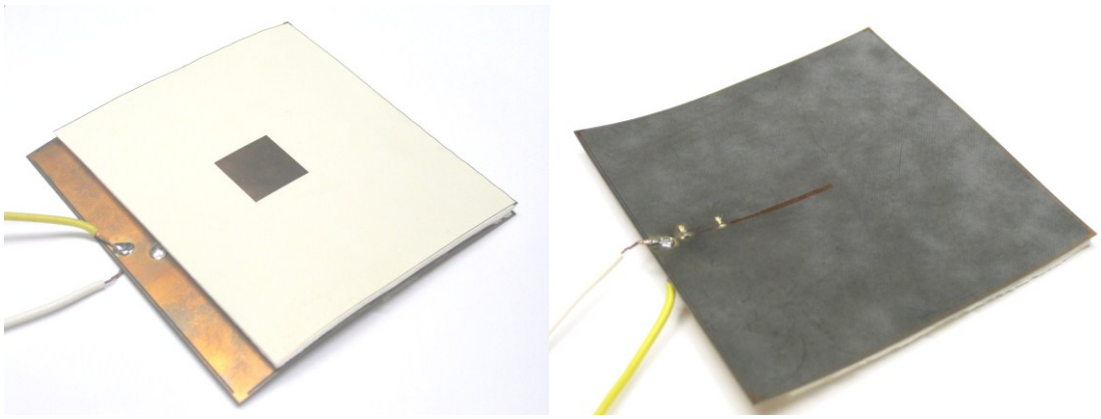


Figure 8.5. Rectenna design (all dimensions in mm).



(a) (b)

Figure 8.6. Photographs of the rectenna. (a) Top view and (b) bottom view.

B. Efficiency Measurement

In order to measure the rectenna efficiency, a known power is transmitted from a standard horn antenna, and the resulting voltage is measured across the rectenna's load. The efficiency η can be calculated by

$$\eta = \frac{P_{DC}}{P_{rec}} = \frac{\left(V_D^2 \big/ R_L \right)}{P_t G_t G_r \left(\lambda_0 \big/ 4\pi r \right)^2} \tag{8.3}$$

where P_{DC} is the DC output power of the rectenna and P_{rec} is the power received by the rectenna. V_D is the voltage measured across the load, and R_L is the load resistance. The received power P_{rec} is calculated using the Friis transmission equation [3] where a standard NARDA horn antenna with a 15 dB gain (G_t) is used to transmit the microwave power (P_t) to the rectenna. The rectenna gain (G_r) is 9.3 dB, λ_0 is the free space

wavelength at 5.8 GHz, and r is the distance between the transmitter and the rectenna. The transmitted power (P_t) can be varied to determine the efficiency of the rectenna at different power densities. The load (R_L) can also be varied to determine the rectenna's optimum load. Power density S_D and effective area A_{er} can be calculated [3] as

$$S_D = \frac{P_t G_t}{4\pi r^2} \tag{8.4}$$

$$A_{er} = \frac{G_r \lambda_0^2}{4\pi} \tag{8.5}$$

which can be used to express received power P_{rec} as

$$P_{rec} = S_D A_{er} \ . \tag{8.6}$$

The measurement setup used for this procedure can be seen in Fig. 8.7. A frequency synthesizer is used to generate the 5.8 GHz signal and an isolator is used to protect the synthesizer. A power amplifier must be used so that the transmitted power is sufficiently high to cover the rectenna's entire operating range. A coupler is used immediately before the transmitting antenna so that the transmitted power can be measured. Most of the power is coupled to the antenna, but a small portion is coupled to a power meter. An attenuator must be used before the power meter to adjust the power to a level which is readable by the power meter. A network analyzer is used to characterize the coupler and determine how much power is coupled to the transmitting antenna and how much is coupled to the power meter. In this way, the power meter can be used to measure the transmitted power. The rectenna is set at a fixed distance in the far field, and the voltage across its load is measured for various transmitted power levels.

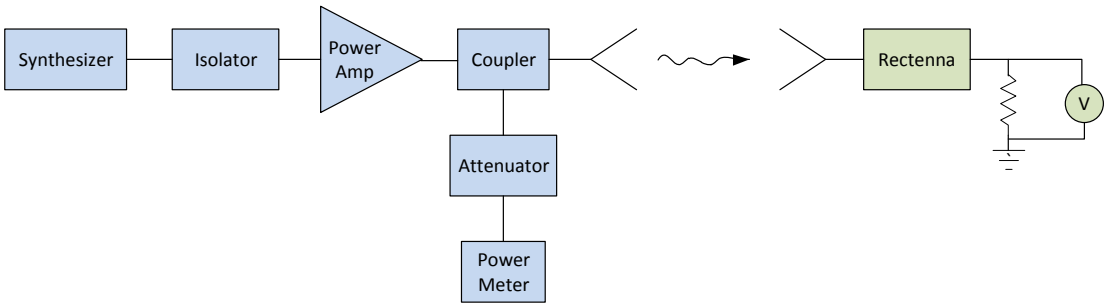


Figure 8.7. Equipment setup for efficiency measurement.

The measured voltage for various loads and power levels can be seen in Fig. 8.8. Higher load values result in higher output voltages. The measured voltage is converted to an efficiency by (8.3), and the results can be seen in Fig. 8.9. A peak efficiency of 73% was measured with a 100 ohm load and occurred at roughly 100 mW received power which is the maximum power rating of the diode. Loads of 80, 90, 110, and 120 ohms were also measured and produced nearly identical results to the 100 ohm load. Other loads ranging from 50 to 500 ohms were measured and showed a predictable decrease in performance the farther the values deviated from 100 ohms. The efficiency of some of the other loads can be seen in the figure.

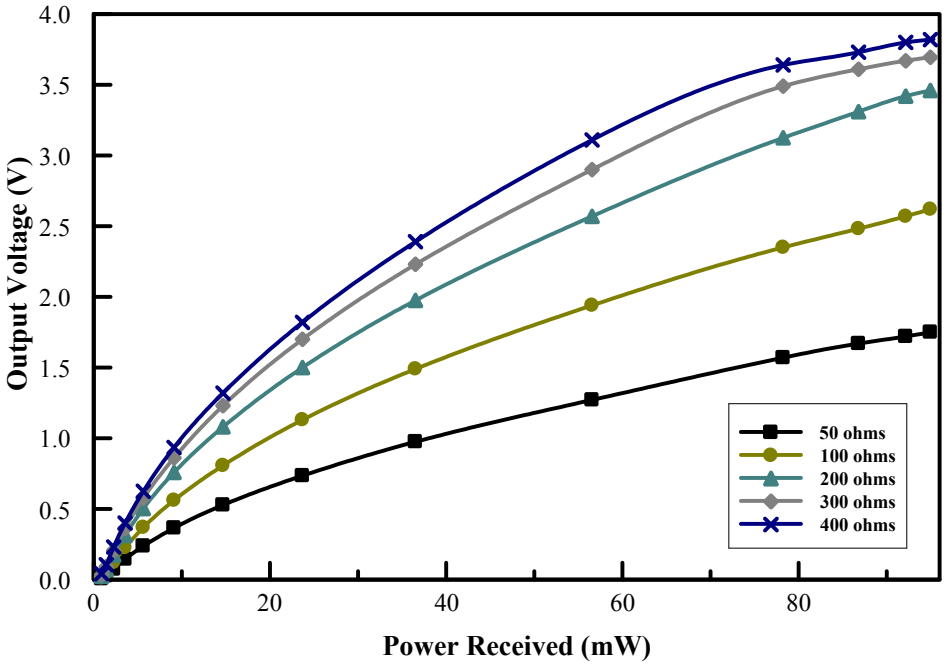


Figure 8.8. Measured rectenna output voltage.

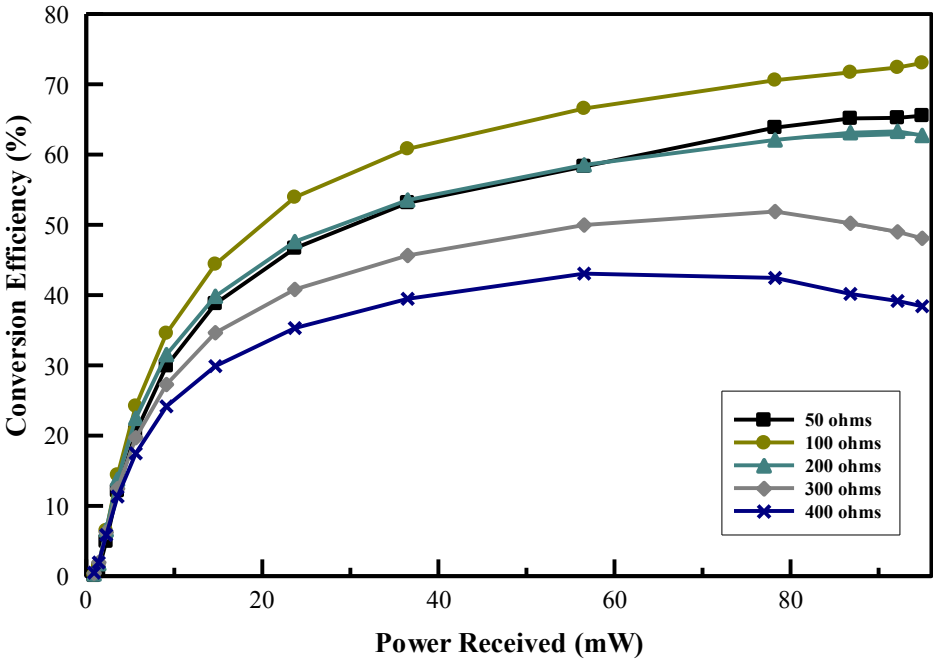


Figure 8.9. Measured RF-to-DC conversion efficiency.

C. Comparison to Model

The rectenna model described in the previous section can be utilized to model this rectenna design, and the simulated results can be compared to the measured results. IE3D [12] can be used to generate S-parameters of the antenna and filter section of the rectenna. These S-parameters can then be used in Agilent ADS [83] along with the diode model, a capacitor, and a resistor to simulate the entire rectenna design. A harmonic balance simulation can then be run with various values for input power and load resistance to verify against the measured results.

The Agilent ADS schematic can be seen in Fig. 8.10. A variable power source and DC-blocking capacitor are used to represent the source of power to the rectenna. Then S-parameters of the filter section (generated from IE3D) are used to represent the rectenna's transmission line up to the quarter-wavelength transformer. The quarter-wavelength transformer and the 120 Ω line (up to the diode) are represented by transmission line components in ADS. The diode is represented by a model component described in the previous section. This is followed by a capacitor with transmission lines on either side and finally a variable resistor. The power source and load can be varied to give output voltages for various power levels and loads.

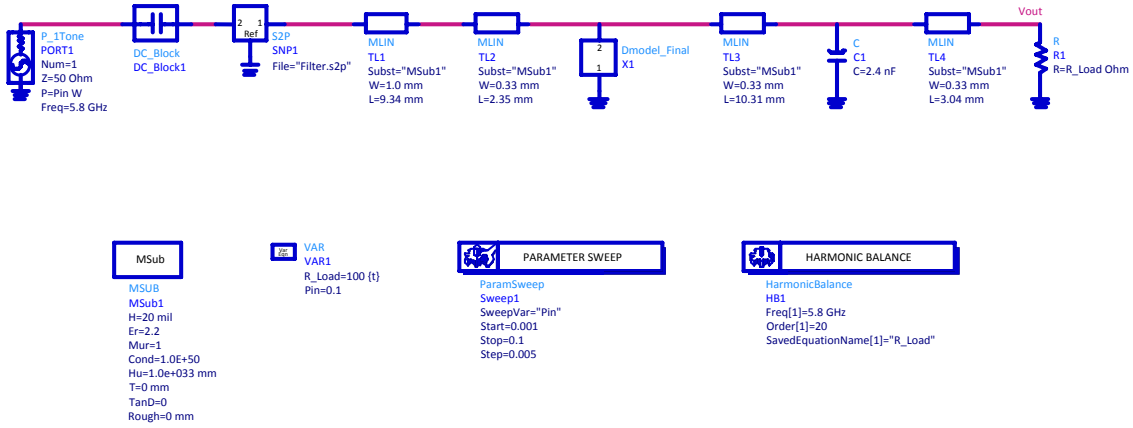


Figure 8.10. Schematic of rectenna simulation in Agilent ADS.

A comparison of measured and simulated output voltages can be seen in Fig. 8.11. The simulated results show a very good match to the measured results. A comparison of measured and simulated efficiency can be seen in Fig. 8.12. Since the output voltage is squared to calculate the efficiency, any error in voltage is magnified in efficiency error. This results in an increased difference between measured and simulated results for efficiency, but the results are still good, and more importantly both the measured and simulated results show an optimum load value of 100 ohms.

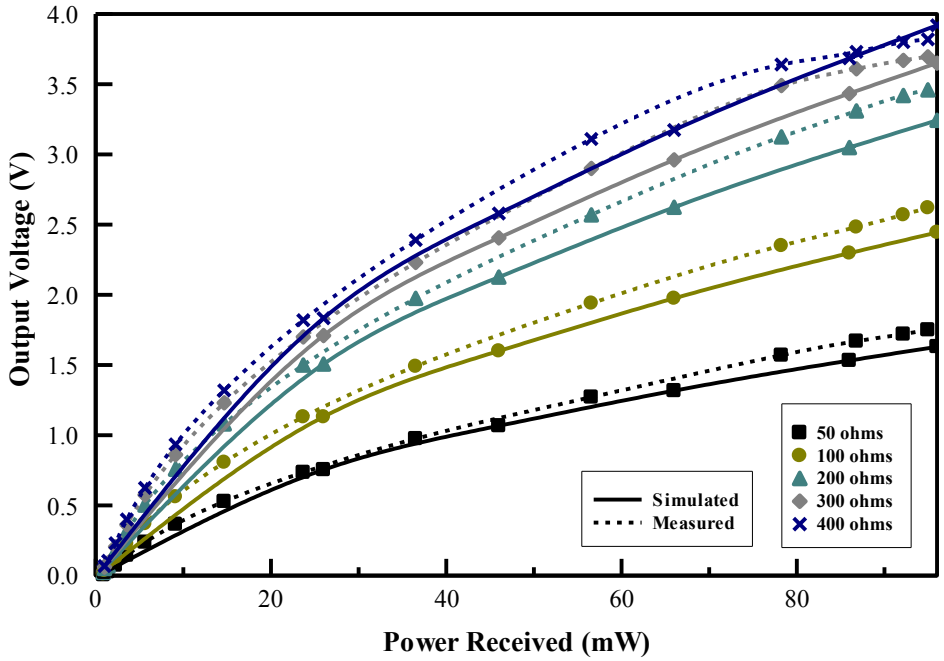


Figure 8.11. Comparison of measured and simulated rectenna output voltage.

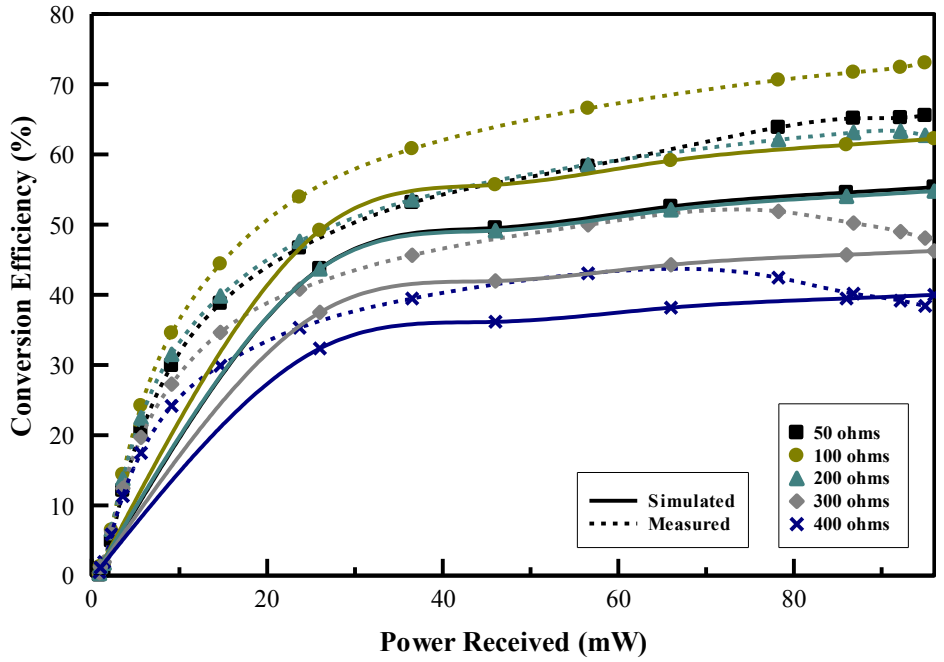


Figure 8.12. Comparison of measured and simulated RF-to-DC conversion efficiency.

D. Conclusion

A 5.8 GHz rectenna was successfully developed with an antenna gain of 9.3 dBi, a second harmonic suppression of 4.93 dB from split ring resonators (SRRs), and a conversion efficiency of 73%. Aperture-coupling was employed to isolate the rectifying operation from the receiving operation. The design was also used to further verify the diode model created for the rectenna in the previous section. The measurements showed a good match to the simulated model.

9. CONCLUSION

A broadband, dual-polarized antenna was designed, fabricated, and measured. The design was tuned using electromagnetic simulation software, but a shift in measured bandwidth necessitated a redesign to account for this unavoidable shift. Feed 1 was measured to have a 42% VSWR bandwidth with f_0=10.0 GHz, and feed 2 was measured to have a 38% VSWR bandwidth with f_0=10.1 GHz. Gain was measured to be 5 – 8 dBi, and the half-power beamwidth was measured to be 50 - 90°. The gain measurements showed a cross-polarization isolation of greater than 20 dB, and the S-parameter isolation between the feeds was measured to be greater than 17 dB. With the exception of the frequency shift, the measured results matched well to the simulation.

This antenna was then used in a broadband, dual-polarized phased array application. A 4x1 antenna array was created from the previously designed single-element which served to increase the gain and allow the antenna pattern to be steered. To feed the antenna array with a uniform amplitude distribution, a 1-to-4 power divider was created, and digital phase-shifters were used to control each element's phase, thereby steering the pattern. A LabVIEW program was created with a simple interface in order to facilitate control of the phase-shifters. After calibrating the phased array at 10 GHz, the array pattern was measured over the target bandwidth for various scan angles. The array performed well over a 3 GHz bandwidth and produced a maximum scan angle of 45°.

Adding more elements to the phased array would greatly improve its performance. The beamwidth could be narrowed and the maximum scan angle increased. Additionally, including controllable amplifiers for each element would permit the use of non-uniform amplitude distributions which could narrow the beamwidth and/or decrease sidelobes. Controllable amplifiers could also increase the calibration accuracy by removing the effect of variable line attenuation on each element. Lastly, using phase-shifters with a more consistent phase shift and insertion loss would also increase the array performance.

Rectennas are a key component of wireless power transmission systems, and the ability to simulate rectenna designs can expedite and improve their development. A diode model which is based on an equivalent circuit was created to serve as an integral part of a complete rectenna model. The model component values were determined from a series of S-parameter measurements (at various bias levels) using a method which systematically isolates the model's voltage-dependent components from its voltage-independent components. The model was implemented in Agilent ADS using symbolically defined devices for both the voltage-dependent current source and the voltage-dependent capacitor, and the model's simulated S-parameters matched well to the measured S-parameters over all bias levels.

An aperture-coupled, high-gain patch rectenna was then designed, fabricated, and measured. This rectenna design was modeled using the previously described method which allowed for a verification of the model's accuracy. The model simulation showed

a good match to the measured results, and both gave the same optimum load value. This demonstrated the model's usefulness for rectenna simulation and design.

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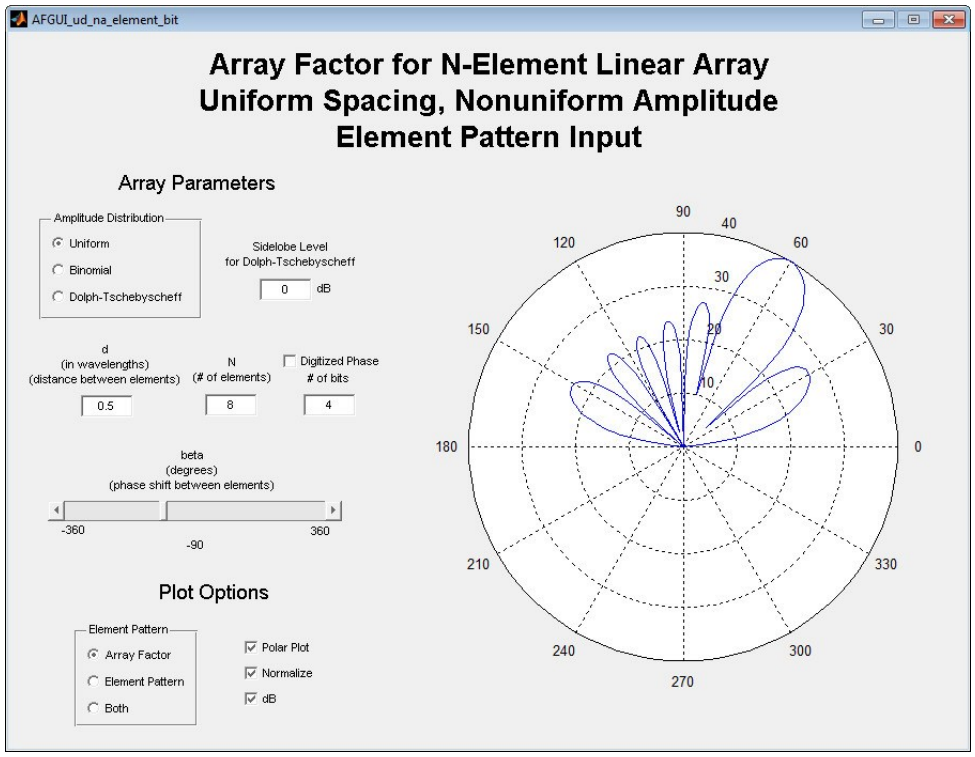
APPENDIX A

MATLAB ARRAY PATTERN VISUALIZATION GUI

This appendix presents a Matlab GUI program which can be used to visualize the effect of different parameters on an array pattern. The phase shift between elements can be varied with a slider to see the effect in real time. Amplitude distribution can be chosen from: Uniform, Binomial, and Dolph-Tschebyscheff; for the Dolph-Tschebyscheff, the sidelobe levels (in dB) can be chosen. The distance between elements and the number of elements can be adjusted. Also, the pattern can be digitized, and the number of bits can be selected.

The plot can be visualized in rectangular or polar coordinates, it can be normalized, and it can be displayed in dB. Also, the array factor, the element pattern, or a combination of the two (real array pattern) can be displayed. The element pattern is input in the Matlab code.



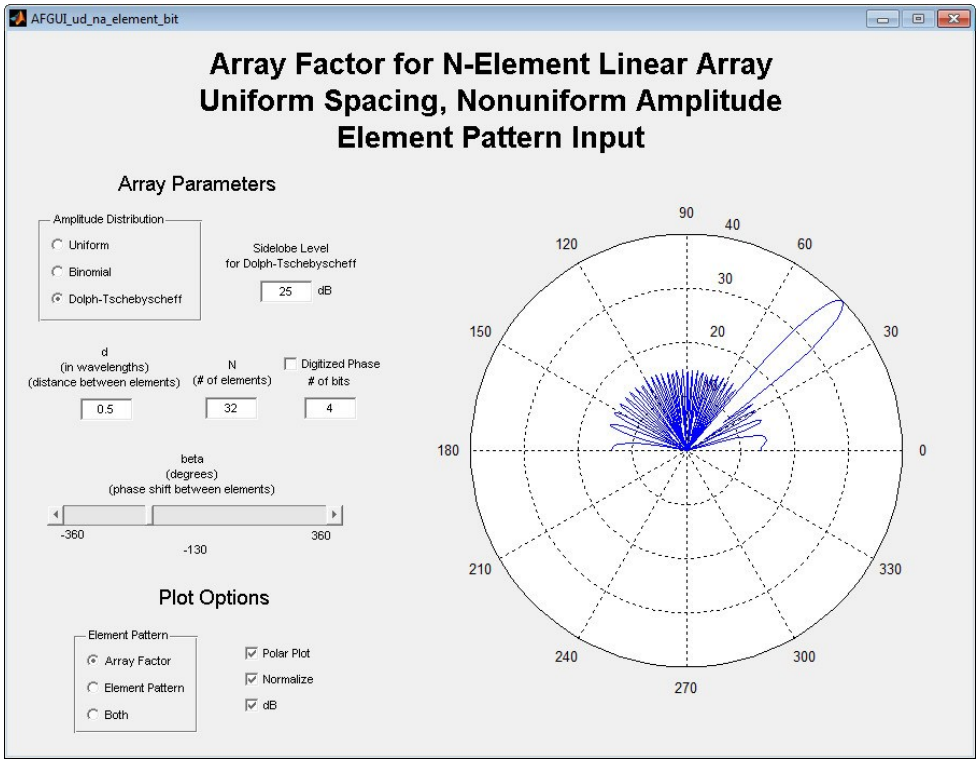


Figure A.1. User interface shown with different settings.

```
function varargout = AFGUI_ud_na_element_bit(varargin)
% AFGUI_UD_NA_ELEMENT_BIT M-file for AFGUI_ud_na_element_bit.fig
%      Jonathan Hansen
%      Array Factor for N-Element Linear Array
%      with Uniform Spacing and Non-uniform Amplitude
%      Allows an Element Pattern to be used and combined with the AF
%
%      Array Factor is calculated with broadside at 90 degrees.
%      IE3D defines a broadside of 0 degrees so this must be shifted.
%
%      Plots the array factor for given parameters.
%         Array Parameters
%           Amplitude Distribution
%              Uniform
%              Binomial
%              Dolph-Tschebyscheff
%         Sidelobe Level for Dolph-Tschebyscheff (in dB)
%         Distance between elements (d) in wavelengths
%         Number of Elements (N)
%         Digitize the phase shift or allow continuous number of bits for digitized
%         Phase shift between elements (beta) (in degrees)
%      Plot Options
%         Element Pattern Options
%            only plot the Array Factor
%            only plot the Element Pattern
%            plot the product of the Array Factor & Element Pattern
%         Polar or Cartesian coordinates
%         Normalize or Non-normalized data
%         dB or Ratio
%
%      AFGUI_UD_NA_ELEMENT_BIT, by itself, creates a new AFGUI_UD_NA_ELEMENT_BIT or raises the existing
%      singleton*.
%
%      H = AFGUI_UD_NA_ELEMENT_BIT returns the handle to a new AFGUI_UD_NA_ELEMENT_BIT or the handle to
%      the existing singleton*.
%
%      AFGUI_UD_NA_ELEMENT_BIT('CALLBACK',hObject,eventData,handles,...) calls the local
%      function named CALLBACK in AFGUI_UD_NA_ELEMENT_BIT.M with the given input arguments.
%
%      AFGUI_UD_NA_ELEMENT_BIT('Property','Value',...) creates a new AFGUI_UD_NA_ELEMENT_BIT or raises the
%      existing singleton*.  Starting from the left, property value pairs are
%      applied to the GUI before AFGUI_ud_na_element_bit_OpeningFunction gets called.  An
%      unrecognized property name or invalid value makes property application
%      stop.  All inputs are passed to AFGUI_ud_na_element_bit_OpeningFcn via varargin.
%
%      *See GUI Options on GUIDE's Tools menu.  Choose "GUI allows only one
%      instance to run (singleton)".
%
% See also: GUIDE, GUIDATA, GUIHANDLES

% Begin initialization code - DO NOT EDIT
gui_Singleton = 1;
gui_State = struct('gui_Name',       mfilename, ...
                   'gui_Singleton',  gui_Singleton, ...
                   'gui_OpeningFcn', @AFGUI_ud_na_element_bit_OpeningFcn, ...
                   'gui_OutputFcn',  @AFGUI_ud_na_element_bit_OutputFcn, ...
                   'gui_LayoutFcn',  [] , ...
                   'gui_Callback',   []);
if nargin && ischar(varargin{1})
    gui_State.gui_Callback = str2func(varargin{1});
end
```

```
if nargout
    [varargout{1:nargout}] = gui_mainfcn(gui_State, varargin{:});
else
    gui_mainfcn(gui_State, varargin{:});
end
% End initialization code - DO NOT EDIT


% --- Executes just before AFGUI_ud_na_element_bit is made visible.
function AFGUI_ud_na_element_bit_OpeningFcn(hObject, eventdata, handles, varargin)
% This function has no output args, see OutputFcn.
% hObject    handle to figure
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)
% varargin   command line arguments to AFGUI_ud_na_element_bit (see VARARGIN)

% Choose default command line output for AFGUI_ud_na_element_bit
handles.output = hObject;

set(handles.AmpSelect_buttongroup,'SelectionChangeFcn',@AmpSelect_buttongroup_SelectionChangeFcn);
set(handles.AFSelect_buttongroup,'SelectionChangeFcn',@AFSelect_buttongroup_SelectionChangeFcn);

% UIWAIT makes AFGUI_ud_na_element_bit wait for user response (see UIRESUME)
% uiwait(handles.figure1);

% Initialize Global Variables
global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
       amp AmpSelect AFSelect SidelobeLeveldB NormalizeF

ElementPatterndB = [-1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761
-1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761
-1992.211761 -1992.211761 -64.79668478 -9.892406686 -7.550849379 -5.933740436 -4.323770356 -2.719428108 -1.173279813
0.275782952 1.60549158 2.805027952 3.870468998 4.801781819 5.600990775 6.27107541 6.81529526 7.236761195 7.538153908
7.721537197 7.788239005 7.738786378 7.572887395 7.289457145 6.886687733 6.362165265 5.713041126 4.936273125
4.028968638 2.988893838 1.815271463 0.510088106 -0.919724216 -2.45876601 -4.078482087 -5.737493634 -7.438667667
-9.871805941 -64.81794103 -1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761
-1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761 -1992.211761
-1992.211761 -1992.211761];
ElementPatterndB = circshift(ElementPatterndB,[0,90]);          % shift by 90 deg to have broadside at 90 deg instead of 0 deg
ElementPattern = 10 .^ (ElementPatterndB ./ 20);                % convert dB to ratio
ElementPattern = ElementPattern ./ max(ElementPattern);         % normalize the Element Pattern

Element_theta = [-3.14159265 -3.05432619 -2.96705973 -2.87979327 -2.7925268 -2.70526034 -2.61799388 -2.53072742
-2.44346095 -2.35619449 -2.26892803 -2.18166156 -2.0943951 -2.00712864 -1.91986218 -1.83259571 -1.74532925 -1.65806279
-1.57079633 -1.48352986 -1.3962634 -1.30899694 -1.22173048 -1.13446401 -1.04719755 -0.95993109 -0.87266463 -0.78539816
-0.6981317 -0.61086524 -0.52359878 -0.43633231 -0.34906585 -0.26179939 -0.17453293 -0.08726646 0 0.08726646 0.17453293
0.26179939 0.34906585 0.43633231 0.52359878 0.61086524 0.6981317 0.78539816 0.87266463 0.95993109 1.04719755
1.13446401 1.22173048 1.30899694 1.3962634 1.48352986 1.57079633 1.65806279 1.74532925 1.83259571 1.91986218
2.00712864 2.0943951 2.18166156 2.26892803 2.35619449 2.44346095 2.53072742 2.61799388 2.70526034 2.7925268
2.87979327 2.96705973 3.05432619 3.14159265];
Element_theta_deg = Element_theta * 180 / pi;   % observation angle (degrees)

theta_deg = 0 : 1 : 180;            % observation angle (degrees) [specify range & resolution]
theta = theta_deg * pi / 180;       % observation angle (radians)
cos_theta = cos(theta);             % pre-compute cos(theta) to save time later.

% Interpolate element pattern to new set of thetas
Element = interp1( Element_theta_deg, ElementPattern, theta_deg );
```

```
small_num = 1e-11;                      %  to prevent divide by zero problem
d = 0.5;                                % distance between elements
N = 10;                                 % number of elements
phase_step_deg = 360 / (2 ^ 50);        % number of steps (degrees) for digitized phase-shifter (phase_step_deg = 360 / 2^#bits)
phase_step = phase_step_deg * pi/180;   % number of steps (radians) for digitized phase-shifter
beta_deg = 0;                           % phase shift between elements (degrees)
beta = beta_deg * pi/180;               % phase shift between elements (radians)
DigitizedPhase = 0;                     % true = digitize the phase of each element with the number of bits specified
PolarPlot = 0;                          % true = display polar plot; false = display cartesian plot
dBPlot = 0;                             % true = plot in dB
NormalizePlot = 0;                      % true = normalize data for the plot
AmpSelect = 1;                          % uniform amplitude distribution
AFSelect = 1;                           % display array factor only
SidelobeLeveldB = 0;                    % Sidelobe Level of 0 dB
compute_amps(hObject, eventdata, handles);
plot_data(hObject, eventdata, handles);
guidata(hObject, handles);  % updates the handles structure


% --- Outputs from this function are returned to the command line.
function varargout = AFGUI_ud_na_element_bit_OutputFcn(hObject, eventdata, handles)
% varargout  cell array for returning output args (see VARARGOUT);
% hObject    handle to figure
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Get default command line output from handles structure
varargout{1} = handles.output;


% --- Executes on slider movement.
function sliderBeta_Callback(hObject, eventdata, handles)
% hObject    handle to sliderBeta (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hints: get(hObject,'Value') returns position of slider
%        get(hObject,'Min') and get(hObject,'Max') to determine range of slider
global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
       amp AmpSelect AFSelect SidelobeLeveldB NormalizeF
beta_deg = get(hObject,'Value');
beta = beta_deg * pi/180;
beta_string = num2str(beta_deg);

% need to convert the answer into String type to display it
set(handles.display_beta,'String',beta_string);
plot_data(hObject, eventdata, handles);

guidata(hObject, handles);  % updates the handles structure


% --- Executes during object creation, after setting all properties.
function sliderBeta_CreateFcn(hObject, eventdata, handles)
% hObject    handle to sliderBeta (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    empty - handles not created until after all CreateFcns called

% Hint: slider controls usually have a light gray background.
if isequal(get(hObject,'BackgroundColor'), get(0,'defaultUicontrolBackgroundColor'))
   set(hObject,'BackgroundColor',[.9 .9 .9]);
end
```

```
function editSL_Callback(hObject, eventdata, handles)
% hObject    handle to editSL (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hints: get(hObject,'String') returns contents of editSL as text
%        str2double(get(hObject,'String')) returns contents of editSL as a double
global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
       amp AmpSelect AFSelect SidelobeLeveldB NormalizeF
SidelobeLeveldB = str2double(get(hObject,'String'));
compute_amps(hObject, eventdata, handles);
if NormalizePlot
   compute_normalize(hObject, eventdata, handles);
end
plot_data(hObject, eventdata, handles);
guidata(hObject, handles);  % updates the handles structure


% --- Executes during object creation, after setting all properties.
function editSL_CreateFcn(hObject, eventdata, handles)
% hObject    handle to editSL (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    empty - handles not created until after all CreateFcns called

% Hint: edit controls usually have a white background on Windows.
%       See ISPC and COMPUTER.
if ispc && isequal(get(hObject,'BackgroundColor'), get(0,'defaultUicontrolBackgroundColor'))
   set(hObject,'BackgroundColor','white');
end


function editN_Callback(hObject, eventdata, handles)
% hObject    handle to editN (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hints: get(hObject,'String') returns contents of editN as text
%        str2double(get(hObject,'String')) returns contents of editN as a double
global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
       amp AmpSelect AFSelect SidelobeLeveldB NormalizeF
N = str2double(get(hObject,'String'));
compute_amps(hObject, eventdata, handles);
if NormalizePlot
   compute_normalize(hObject, eventdata, handles);
end
plot_data(hObject, eventdata, handles);
guidata(hObject, handles);  % updates the handles structure


% --- Executes during object creation, after setting all properties.
function editN_CreateFcn(hObject, eventdata, handles)
% hObject    handle to editN (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    empty - handles not created until after all CreateFcns called

% Hint: edit controls usually have a white background on Windows.
%       See ISPC and COMPUTER.
if ispc && isequal(get(hObject,'BackgroundColor'), get(0,'defaultUicontrolBackgroundColor'))
   set(hObject,'BackgroundColor','white');
end
```

```
function editd_Callback(hObject, eventdata, handles)
% hObject    handle to editd (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hints: get(hObject,'String') returns contents of editd as text
%        str2double(get(hObject,'String')) returns contents of editd as a double
global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
       amp AmpSelect AFSelect SidelobeLeveldB NormalizeF
d = str2double(get(hObject,'String'));
if NormalizePlot
   compute_normalize(hObject, eventdata, handles);
end
plot_data(hObject, eventdata, handles);
guidata(hObject, handles);  % updates the handles structure


% --- Executes during object creation, after setting all properties.
function editd_CreateFcn(hObject, eventdata, handles)
% hObject    handle to editd (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    empty - handles not created until after all CreateFcns called

% Hint: edit controls usually have a white background on Windows.
%       See ISPC and COMPUTER.
if ispc && isequal(get(hObject,'BackgroundColor'), get(0,'defaultUicontrolBackgroundColor'))
   set(hObject,'BackgroundColor','white');
end


function editbits_Callback(hObject, eventdata, handles)
% hObject    handle to editbits (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hints: get(hObject,'String') returns contents of editbits as text
%        str2double(get(hObject,'String')) returns contents of editbits as a double
global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
       amp AmpSelect AFSelect SidelobeLeveldB NormalizeF
phase_step_deg = 360 / (2 ^ str2double(get(hObject,'String')));      % number of steps (degrees) for digitized phase-shifter
                                                                     % (phase_step_deg = 360 / 2^#bits)
phase_step = phase_step_deg * pi/180;                                % number of steps (radians) for digitized phase-shifter
if NormalizePlot
   compute_normalize(hObject, eventdata, handles);
end
plot_data(hObject, eventdata, handles);
guidata(hObject, handles);  % updates the handles structure


% --- Executes during object creation, after setting all properties.
function editbits_CreateFcn(hObject, eventdata, handles)
% hObject    handle to editbits (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    empty - handles not created until after all CreateFcns called

% Hint: edit controls usually have a white background on Windows.
%       See ISPC and COMPUTER.
if ispc && isequal(get(hObject,'BackgroundColor'), get(0,'defaultUicontrolBackgroundColor'))
   set(hObject,'BackgroundColor','white');
end
```

```
% --- Executes on button press in checkboxDigitized.
function checkboxDigitized_Callback(hObject, eventdata, handles)
% hObject    handle to checkboxDigitized (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hint: get(hObject,'Value') returns toggle state of checkboxDigitized
global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
       amp AmpSelect AFSelect SidelobeLeveldB NormalizeF
DigitizedPhase = get(hObject,'Value');
if NormalizePlot
   compute_normalize(hObject, eventdata, handles);
end
plot_data(hObject, eventdata, handles);
guidata(hObject, handles);  % updates the handles structure


% --- Executes on button press in checkboxPolar.
function checkboxPolar_Callback(hObject, eventdata, handles)
% hObject    handle to checkboxPolar (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hint: get(hObject,'Value') returns toggle state of checkboxPolar
global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
       amp AmpSelect AFSelect SidelobeLeveldB NormalizeF
PolarPlot = get(hObject,'Value');
plot_data(hObject, eventdata, handles);
guidata(hObject, handles);  % updates the handles structure


% --- Executes on button press in checkboxdB.
function checkboxdB_Callback(hObject, eventdata, handles)
% hObject    handle to checkboxdB (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hint: get(hObject,'Value') returns toggle state of checkboxdB
global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
       amp AmpSelect AFSelect SidelobeLeveldB NormalizeF
dBPlot = get(hObject,'Value');
plot_data(hObject, eventdata, handles);
guidata(hObject, handles);  % updates the handles structure


% --- Executes on button press in checkboxNormalize.
function checkboxNormalize_Callback(hObject, eventdata, handles)
% hObject    handle to checkboxNormalize (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hint: get(hObject,'Value') returns toggle state of checkboxNormalize
global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
       amp AmpSelect AFSelect SidelobeLeveldB NormalizeF
NormalizePlot = get(hObject,'Value');
if NormalizePlot
   compute_normalize(hObject, eventdata, handles);
end
plot_data(hObject, eventdata, handles);
guidata(hObject, handles);  % updates the handles structure
```

```
% --- Executes on button press in AmpSelect_buttongroup
function AmpSelect_buttongroup_SelectionChangeFcn(hObject, eventdata)

global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
        amp AmpSelect AFSelect SidelobeLeveldB NormalizeF

% retrieve GUI data, i.e. the handles structure
handles = guidata(hObject);

switch get(eventdata.NewValue,'Tag')   % Get Tag of selected object
   case 'radiobuttonUniform'
     % execute this code when radiobuttonUniform is selected
     AmpSelect = 1;  % Uniform

   case 'radiobuttonBinomial'
     % execute this code when radiobuttonBinomial is selected
     AmpSelect = 2;  % Binomial

   otherwise % 'radiobuttonDolph'
     % execute this code when radiobuttonDolph is selected
     AmpSelect = 3;  % Dolph-Tschebyscheff
end

compute_amps(hObject, eventdata, handles);
if NormalizePlot
   compute_normalize(hObject, eventdata, handles);
end
plot_data(hObject, eventdata, handles);

guidata(hObject, handles);  % updates the handles structure


% --- Executes on button press in AFSelect_buttongroup
function AFSelect_buttongroup_SelectionChangeFcn(hObject, eventdata)

global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
        amp AmpSelect AFSelect SidelobeLeveldB NormalizeF

% retrieve GUI data, i.e. the handles structure
handles = guidata(hObject);

switch get(eventdata.NewValue,'Tag')   % Get Tag of selected object
   case 'radiobuttonArrayFactor'
     % execute this code when radiobuttonArrayFactor is selected
     AFSelect = 1;  % Display only Array Factor

   case 'radiobuttonElementPattern'
     % execute this code when radiobuttonElementPattern is selected
     AFSelect = 2;  % Display only Element Pattern

   otherwise  % 'radiobuttonBoth'
     % execute this code when radiobuttonBoth is selected
     AFSelect = 3;  % Display both the Array Factor and the Element Pattern
end

plot_data(hObject, eventdata, handles);

guidata(hObject, handles);  % updates the handles structure
```

```
% --- Compute Amplitudes
function compute_amps(hObject, eventdata, handles)

global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
        amp AmpSelect AFSelect SidelobeLeveldB NormalizeF

% axes(handles.axes1)

switch AmpSelect
   case 2  % Binomial
      P_mat = pascal(N);  % create Pascal's Matrix
      amp = (P_mat( N : N-1 : (N-1)*N+1 ))';  % create row N
   case 3  % Dolph-Tschebyscheff
      amp = chebwin(N,SidelobeLeveldB);
   otherwise  % case 1 - Uniform
      amp = ones(N,1);
end

guidata(hObject, handles);  % updates the handles structure


% --- Compute Normalizing Factor
function compute_normalize(hObject, eventdata, handles)

global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
        amp AmpSelect AFSelect SidelobeLeveldB NormalizeF

% maximum should be at broadside for beta=0;
theta_n = pi/2;
beta_n = 0;

psi = 2*pi * d * cos(theta_n) + beta_n;
NormalizeF = 0;
for n = 1 : N
   NormalizeF = NormalizeF + amp(n)*exp(j*(n-1)*psi);
end

NormalizeF = abs(NormalizeF);

guidata(hObject, handles);  % updates the handles structure


% --- Plot the data
function plot_data(hObject, eventdata, handles)

global beta beta_deg N phase_step d DigitizedPhase PolarPlot dBPlot NormalizePlot theta theta_deg cos_theta Element small_num
        amp AmpSelect AFSelect SidelobeLeveldB NormalizeF

axes(handles.axes1)


% Calculate AF
AF = zeros( 1 , size(theta,2) );     % initialize & reserve memory for AF
pi_d_2 = 2 * pi * d;                 % precompute to save time
if  DigitizedPhase                   % digitize the phase shift of each element
   for k = 1 : size(theta,2)
      psi_beta = pi_d_2 * cos_theta(k);  % psi - beta
      for n = 1 : N
         el_shift = (n-1)*beta;  % phase shift for individual element
         dig_el_shift = round(el_shift / phase_step) * phase_step;  % digitized phase shift for individual element
         AF(k) = AF(k) + amp(n) * exp( j*((n-1)*psi_beta + dig_el_shift) );
```

```
        end
    end
else  % the phase-shifter is continuous
    for k = 1 : size(theta,2)
        psi = pi_d_2 * cos_theta(k) + beta;
        for n = 1 : N
            AF(k) = AF(k) + amp(n)*exp(j*(n-1)*psi);
        end
    end
end
AF = abs(AF);  % take magnitude of complex array factor

% Normalize AF (if required)
if NormalizePlot  % normalize the array factor
    AF = AF ./ NormalizeF;
end

% Multiply AF by Element Pattern if required
switch AFSelect
    case 2  % display only element pattern
        AF = Element;
    case 3  % display array factor and element pattern
        AF = AF .* Element;
end

% Convert AF to dB (if required)
if dBPlot  % plot in dB
    AF = 20*log10(AF);
end

% Plot AF
if PolarPlot  % plot in polar coordinates
    if dBPlot
        if NormalizePlot
            dBshift = 40;               % all data below -dBshift will be set to dBshift
            AF = AF + dBshift;          % shift AF to positive values
            AF = (AF>0) .* AF;          % set negative AF values to 0
        else  % plot is not normalized
            dBshift = min(AF);
            AF = AF - dBshift;          % shift negative dB values for polar plot
        end
    end
    polar(theta,AF);
else  % plot in rectangular coordinates
    plot(theta_deg,AF);
    xlabel('angle (deg)');
    if dBPlot
        if (AFSelect == 1)
            ylabel('Array Factor (dB)');
        elseif (AFSelect == 2)
            ylabel('Element Pattern (dB)');
        else
            ylabel('Array Pattern (dB)');
        end
    else
        if (AFSelect == 1)
            ylabel('Array Factor');
        elseif (AFSelect == 2)
            ylabel('Element Pattern');
        else
            ylabel('Array Pattern');
        end
    end
    set(gca,'XTick',0:30:180)  % set tick marks to 30 deg increments
```

```
    if NormalizePlot
        if dBPlot
            axis([0  180  -40  0])
        else
            axis([0  180  0  1])
        end
    end
end

guidata(hObject, handles);  % updates the handles structure
```

APPENDIX B

LABVIEW PHASE-SHIFTER CONTROLLER

1. PhaseToBinary.vi

This LabVIEW virtual instrument converts a phase shift (in degrees) into a 4-bit number which will control a 4-bit phase-shifter.

Figure B.1. PhaseToBinary.vi icon.

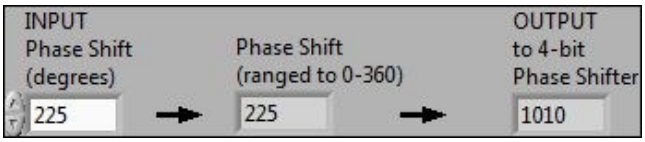


Figure B.2. PhaseToBinary.vi front panel.

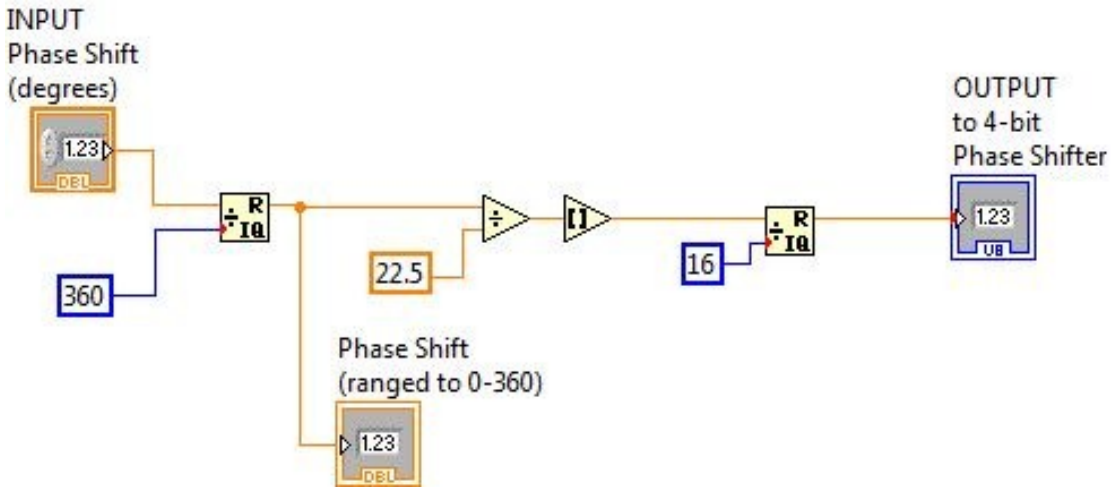


Figure B.3. PhaseToBinary.vi diagram.

2. PhaseToBinaryDB.vi

This LabVIEW virtual instrument converts two phase shifts (in degrees) into an 8-bit number which will control two 4-bit phase-shifters. "Phase Shift 1" will be the LSBs. PhaseToBinary.vi is used in the implementation.

Figure B.4. PhaseToBinaryDB.vi icon.

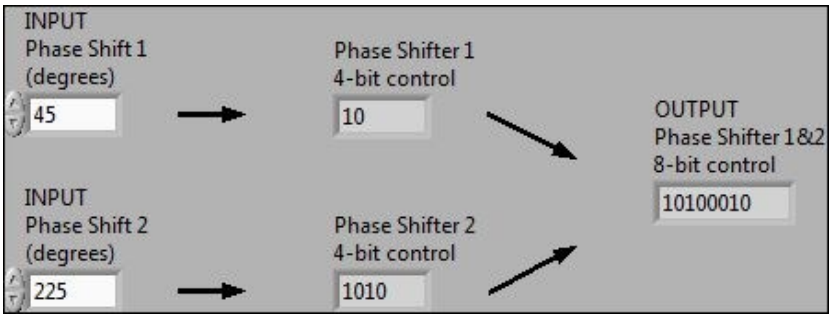


Figure B.5. PhaseToBinaryDB.vi front panel.

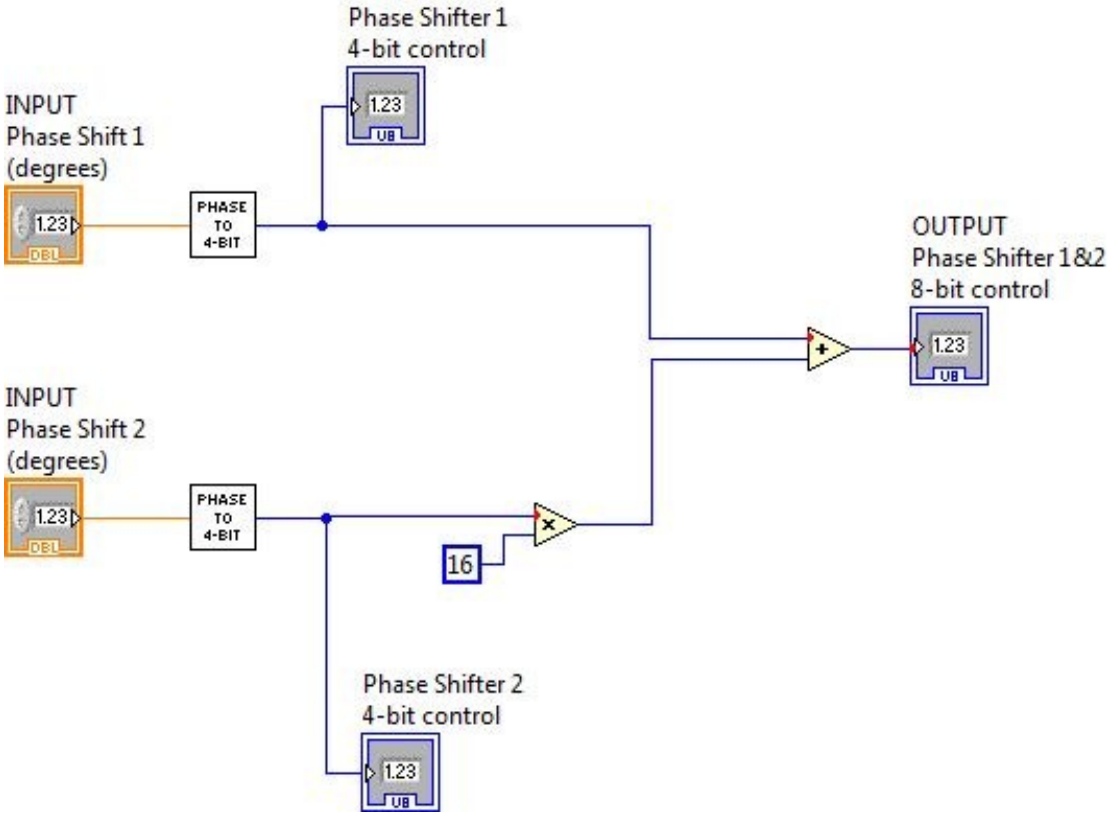


Figure B.6. PhaseToBinaryDB.vi diagram.

3. 4_ElementControl.vi

This LabVIEW virtual instrument controls four 4-bit phase-shifters. Phase-shifters 1 & 2 are on one port, and phase-shifters 3 & 4 are on another. Phase-shifter 1 & 3 are the LSBs of their respective ports, and phase-shifters 2 & 4 are the MSBs of their respective ports. PhaseToBinaryDB.vi is used in the implementation.

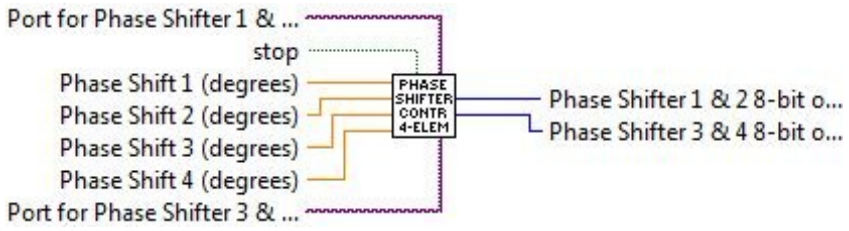


Figure B.7. 4_ElementControl.vi icon.

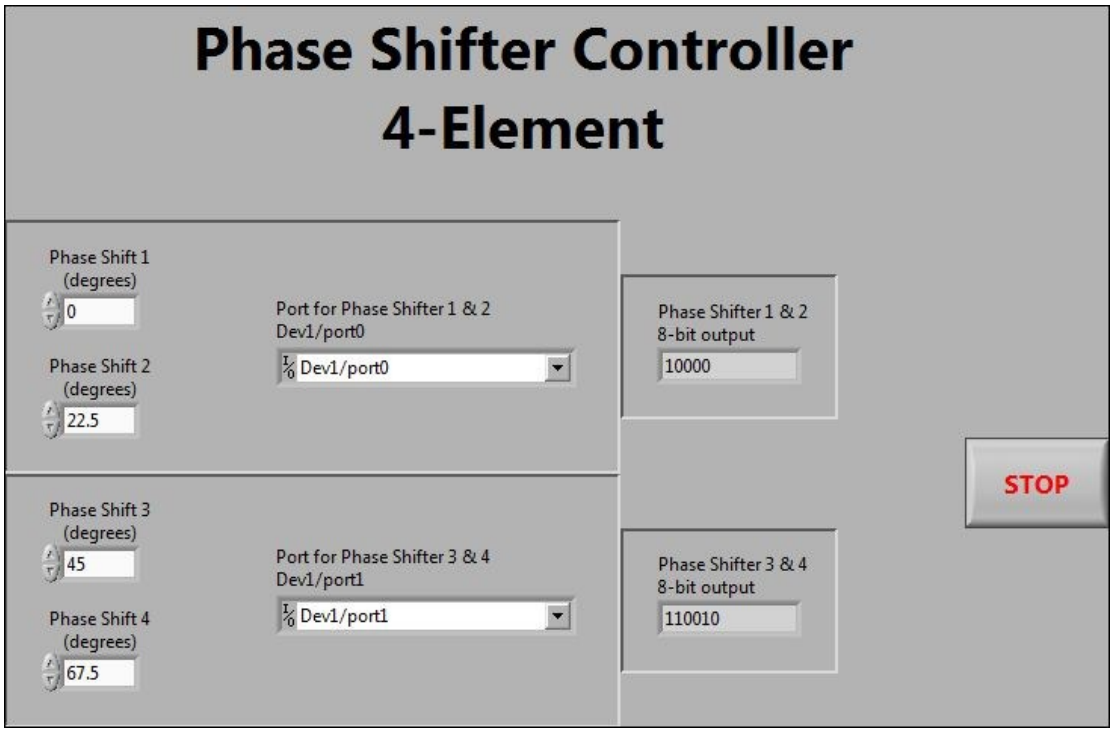


Figure B.8. 4_ElementControl.vi front panel.

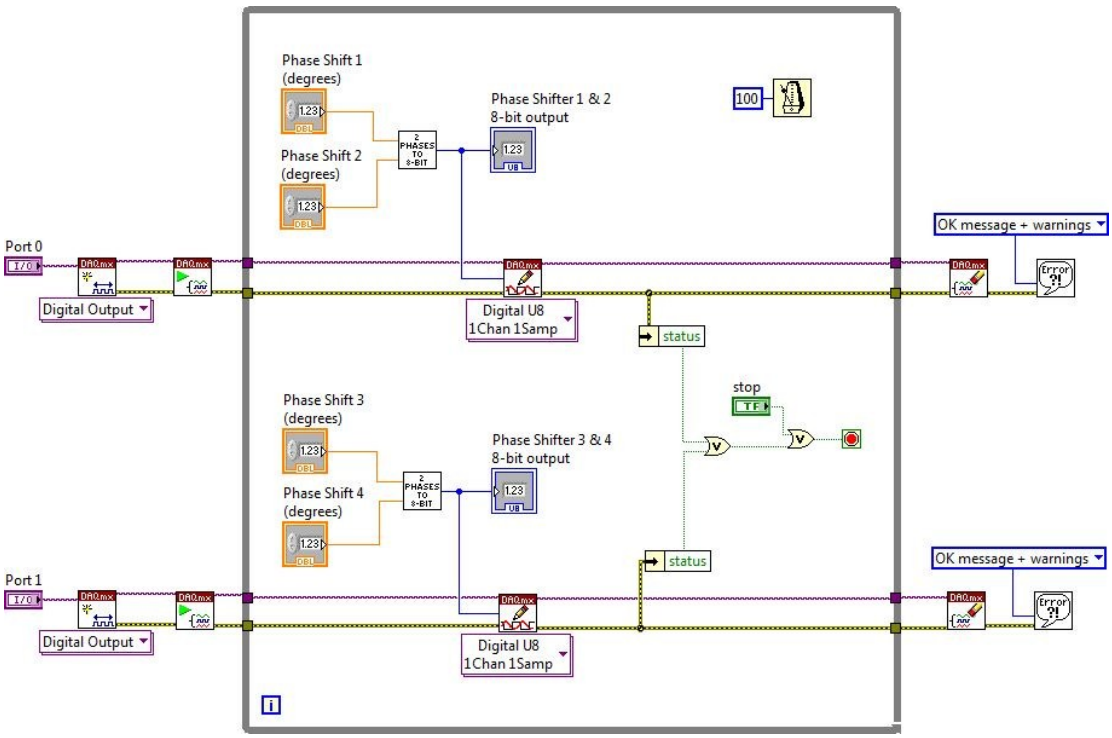


Figure B.9. 4_ElementControl.vi diagram.

APPENDIX C

MATLAB PHASE FINDER

This appendix presents a Matlab program which can be used to find the optimum phase-shifter settings from a calibration table to minimize phase quantization error.

```
% find_phases.m
% Jonathan Hansen
%
% For a list of 4 phase-shifters with 16 phase settings each, find the
% values of each shifter for the desired phase front that will minimize the
% error from the ideal values.

clear all   % removes all variables, globals, functions and MEX links.
clc         % clear the command window


%%%%%  Begin Settings  %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%

% MeasPhase is the measured phase for each setting of each phase-shifter
% rows 1-4 are phase-shifters 1-4
% columns 1-16 are the phase settings for each phase-shifter
MeasPhase = zeros( 4, 16 );  % initialize & reserve memory for MeasPhase
MeasPhase(1,:) = [-100 -82 -59 -45 -28 -14 1 19 36 51 71 88 117 145 174 -151];
MeasPhase(2,:) = [39 54 76 96 129 157 -175 -145 -96 -80 -58 -43 -26 -11 5 24];
MeasPhase(3,:) = [38 52 72 90 115 139 163 -163 -104 -85 -57 -40 -23 -10 5 23];
MeasPhase(4,:) = [38 52 71 89 117 142 172 -153 -96 -79 -55 -40 -22 -10 6 24];

% PhaseSetNames contains the name of each phase setting #1-16
PhaseSetNames = 0 : 22.5 : 337.5;

% ideal angle increase for each phase-shifter
IdealShift = -180*sind(40)

% number of best matches to save
NumOptions = 5;

% number of iterations to test for good matches
num_it = 720;

%%%%%  End Settings  %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
```

```
% vector of best sum of squares errors (each row is the error for a different set)
% best match is at index 1
BestError_ss = 1e50*ones(NumOptions,1);

% matrix of best setting values
% (each row is a set of settings) (each column is a different set)
% best match is at index 1
BestSettings = ones(NumOptions,4);

% i is the ideal phase shift of the first shifter
for i=0:360/num_it:360-360/num_it;  % cycle through 360 degrees

   % ideal phase settings for each phase-shifter #1-4
   if IdealShift == 0
      IdealPhase = [i i i i];
   else
      IdealPhase = i : IdealShift : i+(3*IdealShift);
   end

   for k=1:4            % cycle through each shifter
      for kk=1:16     % cycle through each setting
         % array of differences between the ideal phase & each measured phase for each phase setting
         sub_ang_array(kk) = abs((sub_ang(IdealPhase(k),MeasPhase(k,kk))));
      end
      [error(k),index(k)] = min(sub_ang_array);  % find best phase setting for this try (iteration i)
   end
   error_ss = sum(error.^2);  % sum of squares error over all 4 phase-shifters for this try (iteration i)

   if (error_ss<BestError_ss(NumOptions))  % if this try is better than the last setting in the list
      if sum(ismember(BestSettings,index,'rows'))  % if duplicate setting
         ind = find(ismember(BestSettings,index,'rows'),1);  % index of duplicate setting
         if BestError_ss(ind) > error_ss   % if new error for setting is better than stored error for same setting
            BestError_ss(ind) = error_ss; % store better error
            %  Sort modified setting
            j = ind;
            while (j>1) && (BestError_ss(j) < BestError_ss(j-1))
               [BestError_ss(j) BestError_ss(j-1)] = deal(BestError_ss(j-1),BestError_ss(j));       % swap j & (j-1)
               [BestSettings(j,:) BestSettings(j-1,:)] = deal(BestSettings(j-1,:),BestSettings(j,:));  % swap j & (j-1)
               j = j-1;
            end
         end
      else  % not a duplicate setting
         BestSettings(NumOptions,:) = index;        % replace last stored setting with new one
         BestError_ss(NumOptions) = error_ss;      % replace last stored setting with new one
         % Sort new setting
         j = NumOptions;
         while (j>1) && (BestError_ss(j) < BestError_ss(j-1))
            [BestError_ss(j) BestError_ss(j-1)] = deal(BestError_ss(j-1),BestError_ss(j));      %  swap j & (j-1)
            [BestSettings(j,:) BestSettings(j-1,:)] = deal(BestSettings(j-1,:),BestSettings(j,:)); %  swap j & (j-1)
            j = j-1;
         end
      end
   end

end
```

```
%%%%%  Begin Output  %%%%%%
```

APPENDIX D

MATLAB DIODE MODELING (C_j POLYNOMIAL FIT)

This appendix presents a Matlab program used for diode modeling to fit a polynomial to data points for C_j.

```
% Fit_Cj_Poly.m
% Jonathan Hansen
% fit a polynomial for Cj vs Vj to data points

clear all   % removes all variables, globals, functions and MEX links.
clc         % clear the command window
hold off   % clears the current plot when a new plot is made

% define (Vj,Cj) data points
Vj = [-5, -3, -1, 0, 0.2, 0.4, 0.5895, 0.6477, 0.7026, 0.7249, 0.7333, 0.7430, 0.7502, 0.7561, 0.7632, 0.7695];
Cj = [3.6728e-15, 6.1213e-15, 1.2243e-14, 1.8364e-14, 2.0813e-14, 2.8158e-14, 3.9177e-14, 5.6316e-14, 7.4680e-14,
      1.0774e-13, 1.6528e-13, 3.4280e-13, 5.6316e-13, 7.5905e-13, 8.9372e-13, 1.0651e-12];

Vj = Vj';  % must be column vector for fit
Cj = Cj';  % must be column vector for fit

% fit a curve to the data using linear interpolation between data points
fitobject1 = fit(Vj,Cj,'linearinterp')  % linear interpolation

% plot raw data
plot(Vj,Cj,'o')
hold on  % holds the current plot

% create data points from the linear interpolated curve (fitobject1)
x = -5:0.0001:0.75;
y = feval(fitobject1,x)';

% create a polynomial from the linear interpolated curve data
P20 = polyfit(x,y,20);  % 20 degree polynomial
P25 = polyfit(x,y,25);  % 25 degree polynomial

% plot fit polynomials
xx = (-5: 0.001: 0.9)';    % x data for plotting
f20 = polyval(P20,xx);  % create vector data of fit polynomial
f25 = polyval(P25,xx);  % create vector data of fit polynomial
plot(xx,f20,'r')
plot(xx,f25,'b')
axis([-5  0.9  0  0.2e-11])
```

APPENDIX E

MATLAB DIODE MODELING (C_j EXPONENTIAL FIT)

This appendix presents a Matlab program used for diode modeling to fit an exponential equation to data points for C_j. The function Cjerror.m is called from Fit_Cj_Exp.m to calculate the sum of squares error for a given set of parameters.

```
% Fit_Cj_Exp.m
% Jonathan Hansen
% fit an exponential curve for Cj vs Vj to data points
%
% Cr(v) = Co * sqrt(Vo/(Vo-v))                              , v < alpha*Vo
% Cf(v) = Cr(alpha*Vo) + Cr'(alpha*Vo) * (v-alpha*Vo)   , v > alpha*Vo
%
% adjust Co, alpha, & Vo for least squares error between equation & data

clear all   % removes all variables, globals, functions and MEX links.
clc         % clear the command window

% define (Vj,Cj) data points
Vj = [-5, -3, -1, 0, 0.2, 0.4, 0.5895, 0.6477, 0.7026, 0.7249, 0.7333, 0.7430, 0.7502, 0.7561, 0.7632, 0.7695];
Cj = [3.6728e-15, 6.1213e-15, 1.2243e-14, 1.8364e-14, 2.0813e-14, 2.8158e-14, 3.9177e-14, 5.6316e-14, 7.4680e-14,
      1.0774e-13, 1.6528e-13, 3.4280e-13, 5.6316e-13, 7.5905e-13, 8.9372e-13, 1.0651e-12];

% define parameters for fminsearch
MaxFunEvals = 100000;   % Maximum number of function evals allowed (default = 200*numberofvariables = 600)
MaxIter     = 100000;   % Maximum number of iterations allowed (default = 200*numberofvariables = 600)
TolX        = 1e-15;    % Termination tolerance on x (default = 1e-4)
TolFun      = 1e-15;    % Termination tolerance on the function value  (default = 1e-4)

% initial values for parameters
Co = 1.452e-14;
Co = Co * 1e14;  % scale Co by 1e-14 (works better for fminsearch)
Vo = 0.7214;
alpha = 0.9907;

% set options for fminsearch
options = optimset('MaxFunEvals', MaxFunEvals, 'MaxIter', MaxIter, 'TolX', TolX, 'TolFun', TolFun);

% search for minimized Cj error as defined in Cjerror
[x,fval,exitflag,output] = fminsearch('Cjerror',[Co,Vo,alpha],options);
CO    = x(1) * 1e-14  % Co needs to be scaled back to correct value
VO    = x(2)
ALPHA = x(3)
fval        % sum of squares error
exitflag    % 1 = converged, 0 = max iter reached, -1 = term by output func
output      % summary
```

```
% plot best fit curve & raw data
x = -5 : 0.001 : ALPHA*VO;     % x for reverse bias
xx = ALPHA*VO : 0.001: 0.9;   % x for forward bias
Cr = CO .* sqrt(VO./(VO-x));
diff_Cr = 0.5*CO/VO*(1-ALPHA)^(-3/2);   % derivative of Cr
Cf = CO*sqrt(VO/(VO-VO*ALPHA)) + diff_Cr * (xx-ALPHA*VO);
plot(Vj,Cj,'o', x,Cr,'m', xx,Cf,'g');
axis([0  1  0  1.8e-12])


%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%


function error = Cjerror(x)
% Jonathan Hansen
% calculates the sum of squares error between measured data and a function
%
% Cr(v) = Co * sqrt(Vo/(Vo-v))                                  , v < alpha*Vo
% Cf(v) = Cr(alpha*Vo) + Cr'(alpha*Vo) * (v-alpha*Vo)   , v > alpha*Vo

% extract function parameters from input variable
Co = x(1);
Vo = x(2);
alpha = x(3);

% raw data
Vj = [-5, -3, -1, 0, 0.2, 0.4, 0.5895, 0.6477, 0.7026, 0.7249, 0.7333, 0.7430, 0.7502, 0.7561, 0.7632, 0.7695];
Cj = [3.6728e-15, 6.1213e-15, 1.2243e-14, 1.8364e-14, 2.0813e-14, 2.8158e-14, 3.9177e-14, 5.6316e-14, 7.4680e-14,
      1.0774e-13, 1.6528e-13, 3.4280e-13, 5.6316e-13, 7.5905e-13, 8.9372e-13, 1.0651e-12];

xdata = Vj;
ydata = Cj * 1e14;   % scale Cj by 1e14

% large error for unacceptable input values
LARGE_ERROR = 1e50;

if (alpha >=1 || Co <= 0 || Vo <=0)  % unacceptable values
   error = LARGE_ERROR;
else
   Cr_at_Vo_alpha = Co * sqrt( 1 / (1-alpha) );  % Cr(Vo*alpha)
   Vo_alpha = Vo * alpha;

   error = 0;  % initialize sum of squares error
   for k = 1 : size(xdata,2)
      if xdata(k) < Vo_alpha  % reverse bias
         Cr = Co * sqrt( Vo / (Vo-xdata(k)) );
         error = error + (Cr-ydata(k))^2;
      else  % forward bias
         diff_Cr = 0.5*Co/Vo*(1-alpha)^(-3/2);  % derivative of Cr
         Cf = Cr_at_Vo_alpha + diff_Cr * (xdata(k)-Vo_alpha);
         error = error + (Cf-ydata(k))^2;
      end
   end
end
```

VITA

Jonathan Noel Hansen received his Bachelor of Science degree in electrical engineering from Texas A&M University, College Station, TX in 2001 where he graduated *summa cum laude*. He then received his Ph.D degree in electrical engineering from Texas A&M University in 2011.

From 2001 to 2008, he was with Lockheed Martin Aeronautics, Fort Worth, TX, where he supported avionics software design, implementation, and validation for F-35 sensor fusion. He also provided research and development expertise for revolutionary technology, internal projects.

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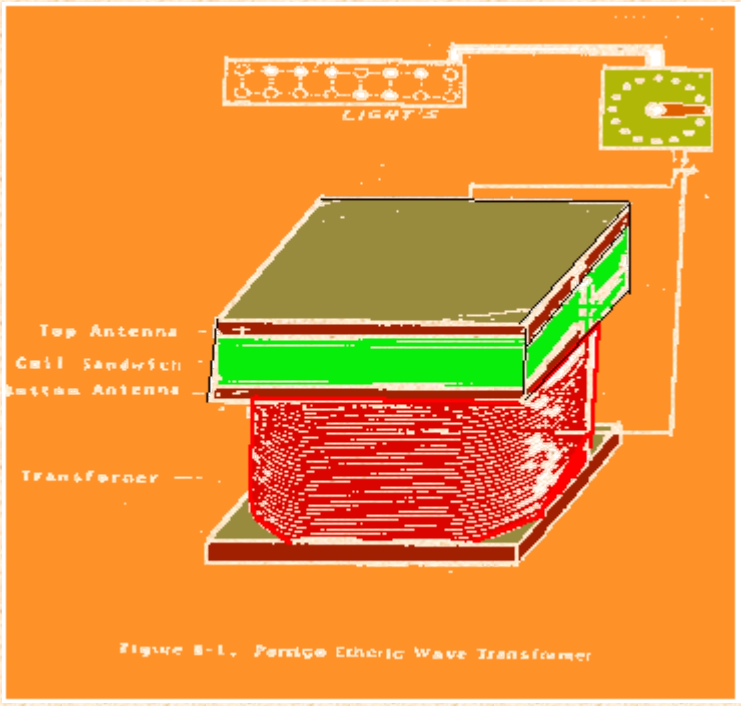


Figure 8-1. Perrigo Etheric Wave Transformer

METHOD AND APPARRATUS FOR ACCUMULATING ELECTRIC ENERGY AND TRANSFORMING ETHER ELECTRIC ENERGY

Drawings by Geoff Egel ,Content Author unknown

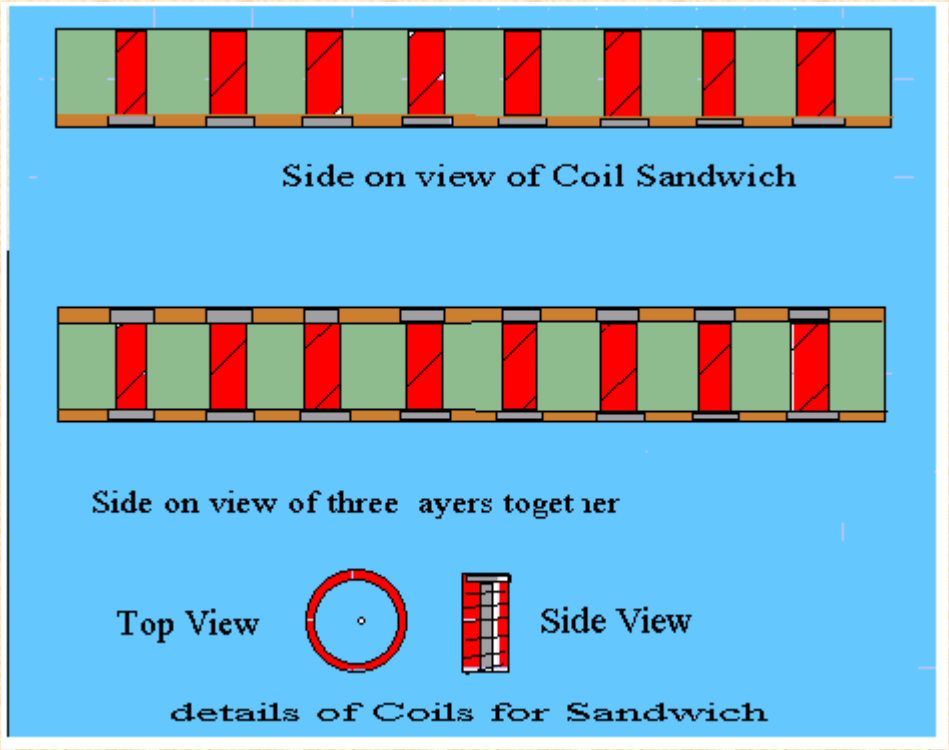
In 1926. Harry E. Perrico claimed to have discovered a method to tap the energy from atmospherics. He even had a car which he converted to run on electricity. that was generated in his ether wave accumulator".

He applied for a patent December 31 1925 with serial number 78,715 being assigned. Perrigo's patent application Is considered a "classic" In patent law and Is listed under the classification of Perpetual Motion Machines and other Impossible Inventions.

Despite the Patent Office's attitude towards Perrico's invention, there were a number of reputable people who claimed to have witnessed his device In action producing useable electric power.

The existence of electromagnetic radiation, the modem term, or electric waves In the ether as it used to be called was known by Hertz, a research scientist who discovered the photoelectric effect

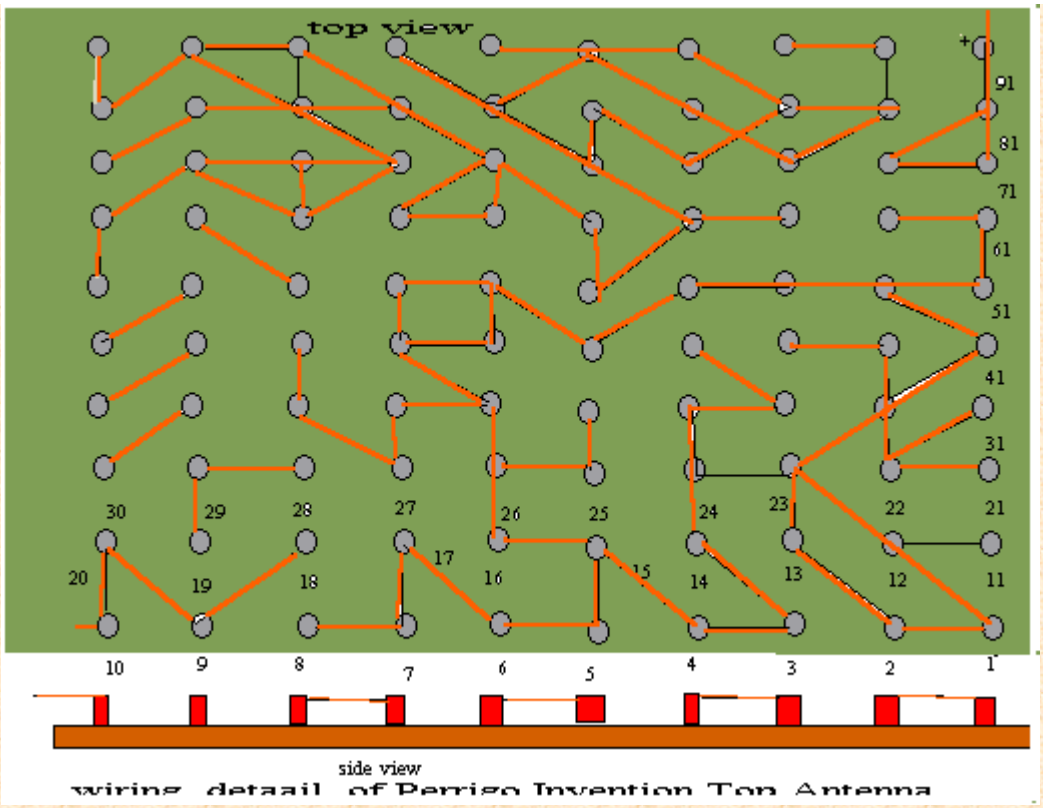
In 1887. Atmosphierics, an electrical disturbance in the atmosphere, were known to produce noises In the early radio telegraph stations, some being strong enough to drown out the received signal.



> Perrigo deduced that here was a possible source of electrical power.

All that was needed was a method of transforming the existing radiation Into useable energy. Hie claimed to have developed a mechanism to intercept and collect from the general ether field electric wave energy and to transform it into useable electromotive force.

The basic method he used was an antenna arrangement which collected and resisted the Incoming energy raising It to a high enough current level where it could be run through a special electrical transformer to further intensify the available power.

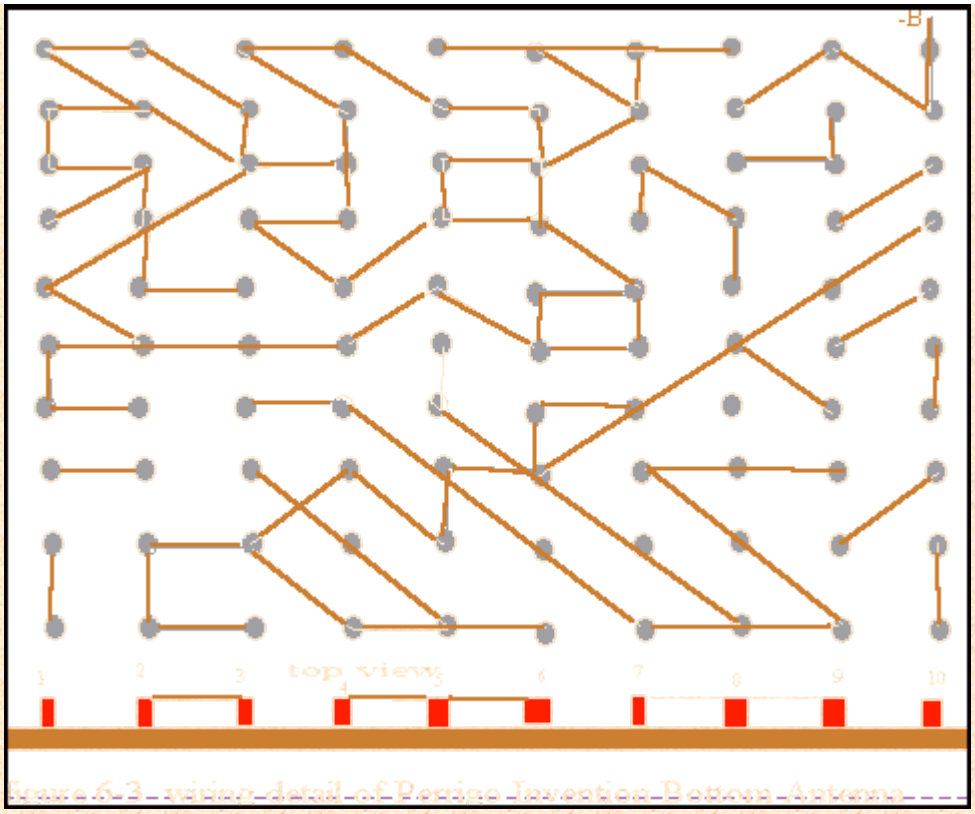


wiring detaail of Perrigo Invention Top Antenna

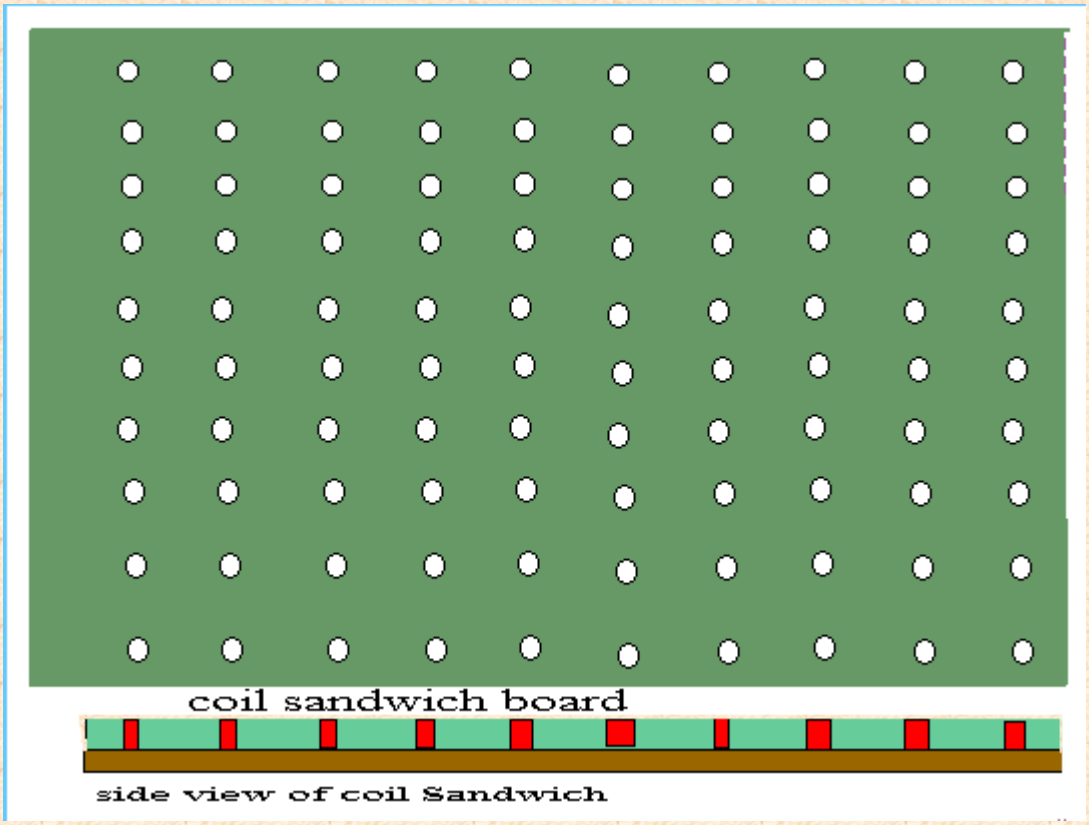
Perrigo's antenna was derived from his experiments with various wire shapes, sizes and arrangements. One of his more successful attempts was to partially pound 100 roofing nails into a board a 10 x 10 array and wrap very fine wire around each nail,making it a small electromagnet.

Then by trial and error approach he connected the ends of the electromagnets to other nails in such a way that there was a maximum voltage between the wire and the nail.. His patent application mirrored this electrical connection scheme in a more refined electromechanical approach.

Two accumulator plates were made with 100 round protruding knobs in a square 10 X 10 array



The accumulator plates were then sandwiched together with an insulator material between them. The insulator had 100 holes matching the protrusions on the plates. Placed in each hole was a special coil wrapped around a bundle of wire . Once the accumulator plates were sandwiched ,a measureable electrical voltage existed between points A and B on plates one and two respectively.



A very complicated transformer was attached to these two points.

The plates were set on top of the transformer and Perrigo claimed this arrangement enhanced the energy accumulation process.

I have no idea of what materials the plate or protrusions were made The patent drawing would lead one to believe they are the same material.

It could be a metal or a nonconductor. such as wood or a combimation. The protrusions were connected by the same wiring scheme

Previously mentioned for the roofing nail model. The connections were different for the two plates.

(See diagrams for the plates connections

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CENTRE TECNOLÒGIC DE TELECOMUNICACIONS DE CATALUNYA

Ultra-Wideband (UWB) rectenna design for Electromagnetic Energy Harvesting

by
César Meneses Ghiglino

A thesis submitted in partial fulfillment for the
degree of Engineer.

In the
Escola Técnica Superior d'Enginyeria de Telecomunicació de Barcelona
Departament de Teoria del Senyal i Comunicacions

October 2010

UNIVERSITAT POLITÉCNICA DE CATALUNYA
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Abstract

Escola Técnica Superior d'Enginyeria de Telecomunicació de Barcelona
Departament de Teoria del Senyal i Comunicacions

degree of Engineer

by César Meneses Ghiglino

This work focuses on designing, fabricating, measuring and testing each component of a Rectenna. A rectenna consists of an antenna and a rectifier circuit that is optimized for incoming signals of low power density. This rectenna is used to harvest electric energy from the RF signals that have been radiated at:

- GSM-850
- GSM-900 (downlink: *935-960MHz*)
- GSM-1800 (downlink: *1805-1880MHz*)
- ISM band centered in 2.45 GHz.

This work contains antenna design techniques using Ansoft HFSS software and methods to simulate rectennas using Harmonic Balance and electromagnetic full-wave Momentum with the Agilent Advanced Design Software (ADS2008).
For the antenna fabrication it was used a LPKF Milling machine. And for measurements a Vector Network Analyzer (VNA), spectral analyzer, analog signal generator, multimeter, and anechoic chamber were used.

Agradecimientos

En las siguientes líneas, quiero agradecer a aquellas personas, a las que considero más importantes en esta etapa, sin el apoyo emocional, material, etc. por su parte, este trabajo hubiera sido más duro.

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Chapter 1

Introduction

Wireless multimedia systems are receiving increasing research and application interests. But improvements are still required to provide higher data-rate links, for instance, the transmission of video signals. Therefore, ultra-wideband (UWB) communication systems are currently under investigation and the design of a compact wideband antenna is very essential. To overcome the inherently narrow bandwidth of microstrip antennas, various techniques have been developed to cover the entire UWB bandwidth, such as L-/F-shaped probe to feed the patch, triangular patch, U-/V-slot monopoles, among others.

And, why don't we use the same antenna to collect the energy that is being radiated at several frequencies? Clearly it is much more effective to harvest the energy of several services at the same time than collecting only one service energy; this is why the idea of an UWB antenna for harvesting energy is entirely feasible.

The meaning of Energy Harvesting (also called energy scavenging or power harvesting), is the process by which energy from different sources is captured and stored. Generally, this definition applies when we talk about autonomous devices that require a low amount of energy to function.

Currently, energy harvesters do not provide sufficient amount of power to produce mechanical movements or temperature changes (cooks, refrigerators, etc) because there

aren't technologies that capture energy with great efficiency. But these technologies do provide the amount of energy needed for low-power devices that can operate autonomously.

Another advantage of this type of technology is that, unlike the production of large-scale power, we can consider that the energy source is free if you take into account the electromagnetic energy of transmitting mobile stations and radio and TV broadcasting antennas.

The use of batteries has two disadvantages: the lifetime of the batteries is very limited even for low-power batteries, requiring impractical periodical battery replacement, the use of commercial batteries usually overkills the power requirements for uW sensor nodes, adding size and weight while creating the problem of environmental pollution due to the deposition of these batteries, as well as increases significantly the cost overhead of disposable nodes.

This work focuses on incident low-power density; designing, measuring and testing a Rectenna to harvest electric energy from the RF signals that have been radiated by public communications systems (GSM-900 and GSM-1800) and the 2.4 GHz ISM band; also the work is motivated by two types of applications: powering low-power sensor networks and RF energy recycling

1.1 Overview

In this thesis, the work involves the design, simulation, fabrication and measurement of each of the components of a rectenna (rectifying antenna circuit) has been divided into the following chapters.

- Chapter 1 is this overview.
- Chapter 2 reviews some of the history of microwave power transmission from its beginnings with the empirical work of Hertz, to Tesla's experiments and describing the evolution of power transmission in free space. Subsequently, it describes how the approaches of some projects, together with studies funded by various companies, encourage the evolution of technologies in the field of microwave power transmission. After the history, an overview of some low power sources for energy harvesting is shown, then we emphasize in the RF-EM energy harvesting and a feasibility study for our application is shown.
- Chapter 3 first describes the main parameters of an antenna for its design. A brief background about UWB antennas and microstrip antennas is introduced. The introduction of the design software (Ansoft HFSS 11) and a brief description of the

simulation method used for the antennas will be shown as well. Then a description of the design procedure and simulations of our base-antenna, which bandwidth range from 850 MHz to 6 GHz is explained. Thereafter, the miniaturization techniques that we applied to the base-antenna and the subsequent building procedures of the “size-improved" antenna will be shown. Finally, some pictures and measures will be explained.

- Chapter 4 describes everything about the rectifier circuit. It begins by describing the operation of a half-wave rectifier and an introduction to the Schottky diode is made, then a brief review of the efficiency of power conversion from RF to DC is shown. For rectifier circuit simulations, the software used was the Agilent ADS2008; this chapter also explains each of the tools used to perform these simulations. For a good overall efficiency, several rectifier circuits were tested, in this chapter, the simulations of each of them and their schemes are detailed. In order to avoid the soldering of a high number of components that may affect the overall efficiency, a transformation from LC components to Microstrip Lines, and their simulations, will be explained. Finally the results of the measurements are shown.
- Chapter 5 shows some possible future applications of rectennas as well as some new substrates materials for antennas that are currently being studied and tested. Finally the conclusions are shown.

Chapter 2

State of the Art

2.1 History of Power Transmission by Radio Waves.

When we talk about power transmission by radio waves, we mean a complex process that can be divided into three stages: 1) to transform the DC power in RF power, 2) RF transmit this RF power through a wireless medium from one point to another; and 3) turn back the received RF power to DC power. Then, we can conclude that the overall system efficiency is directly dependent on the efficiency of each of these stages. The modern history of power transmission studies, in many respects, has focused on improving and developing the elements of the transmitting and receiving ends of this system, focusing on high efficiency, low costs, reliability and low mass.

2.1.1 The Early History

Power transmission dates back to the early work of Hertz [1], He wasn't only the first who showed the propagation of electromagnetic waves in free space, he was also the first to experiment with parabolic reflectors at both the transmitter and receiver ends.

At the end of the 19th century, Nikola Tesla, a genius in the area of generation and power transmission became interested in their studies applied to power transmission from one place to another without wires. His first attempt was carried out in 1899 in Colorado, USA,

with the support of the Colorado Springs Electricity Company. He built a giant coil over which rose a mast of 70 meters with a copper sphere at the end. However, there is no record of how many energy was radiated to the space and whether there was any collected energy at any other place. From today's perspective, we can say that Tesla's attempts of efficient power transmission were ahead of the developed technology of this time. 30 years later, in 1934, H. V. Noble, this time in a laboratory, using two 100 MHz dipoles separated by 8 meters, collected several hundred of watts. With this experiment, he laid the basis for power transfer demonstrations.

The main cause of the lack of interest to develop this area was that several experts had determined that to achieve high efficiencies, concentrations of electromagnetic energy in a narrow band were needed. During the first 35 years of the 20th century, there were no devices that have these capabilities, but in the late 30's, the development of the velocity-modulated beam tube (Klystron tube) and the microwave cavity magnetron, allowed the generation of microwave power dedicated to this new technology.

2.1.2 The Modern History of Free-Space Power Transmission.

The modern history of power transmission, as far as microwaves are concerned, not only includes the development of technologies for microwave power transmission (MPT), the approach also includes various applications, whose achievements contributed to the development of new ideas and new technologies. This history of free-space power transmission will be divided in the early period beginning in 1958 and the period beginning in 1977 with the assessment study, by the DOE / NASA, of the concept of the solar-power satellite.

2.1.2.1 The Early Beginning

In the late 50's, several developments, made the idea of transmitting large amounts of power with high efficiency a reasonable concept, among these developments was the growing need to be able to communicate over long distances with line of sight. This could be solved with some kind of platform located at high altitude. Subsequently, this was solved with a satellite, but back then, a microwave powered vehicle was a logical approach.

Raytheon Company proposed a platform supported by a large helicopter, but did not get the necessary support of the Defense Department; however, they encouraged the development of technologies necessary to make it feasible. Among the technologies that were needed to develop were the high-power microwave tubes required for the transmitter and the technology to directly convert microwave-dc power to drive the motors of helicopter rotors. Between the years 1960-63 were developed some solutions to these problems (the Amplitron and thermionic rectifier respectively) and they did some demonstrations to

receive funds, but some flaws (as the high directivity of the receiving antennas attached to the helicopter and the short-life of thermionic diode) made that the project couldn't be successfully finalized. Due to the low collection efficiency of energy and the directivity of the horns, the concept of rectenna was conceived. This solution consisted in to attach the rectifiers in half-wave dipoles and put a reflector panel behind them. With the application of this rectenna concept the helicopter project (Fig 2.1), the first of July 1964, the first flight of a heavier than air vehicle supported solely by power received from a microwave beam was made.

2.1.2.2 Sponsorship by Marshall Space Flight Center (MSFC)

In 1970, due to the emphasis on developing a free-space power transmission system for a set of physically separate satellites, the MSFC, contracted Raytheon for a system study to be carried out in 1970. In this experiment conducted in 1970, the overall system efficiency obtained was 26%; over the next four years, with the aim of improving the overall system efficiency some other developments were made like a better antenna design, the develop of better ways to launch the microwave beam, the develop for better tools to measure the overall efficiency and better analytical tools for a better understanding of the operation of the rectenna.

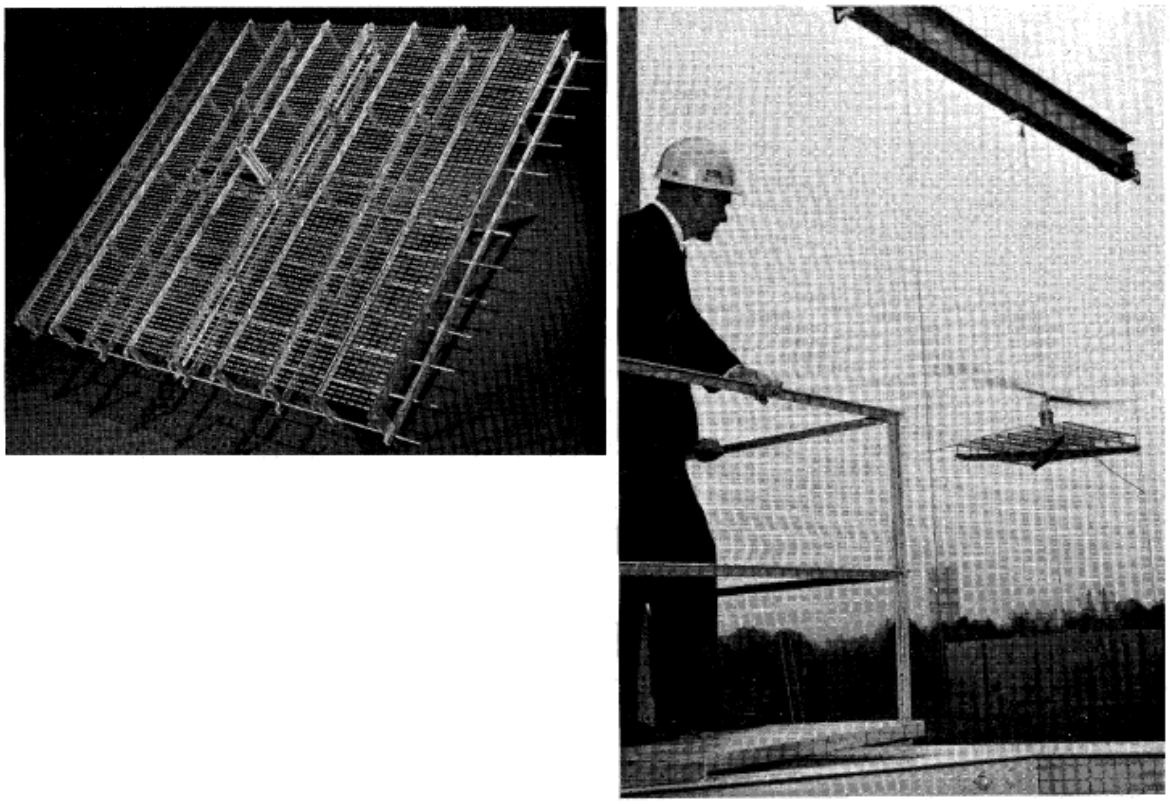


Fig 2.1.1 (left) shows the rectenna made to fly the first helicopter powered by microwaves and (right) the complete system in a test flight of 10 hours, held in October 1964

In 1971, the introduction of the gallium arsenide Schottky-barrier diode with an improved efficiency and greater power handling capacity that the diode that was previously used. This sponsorship program culminated with the construction of a laboratory power transmission system with all the technology that had been developed during the last years. In this experiment, the efficiency obtained was 48% +- 2%.

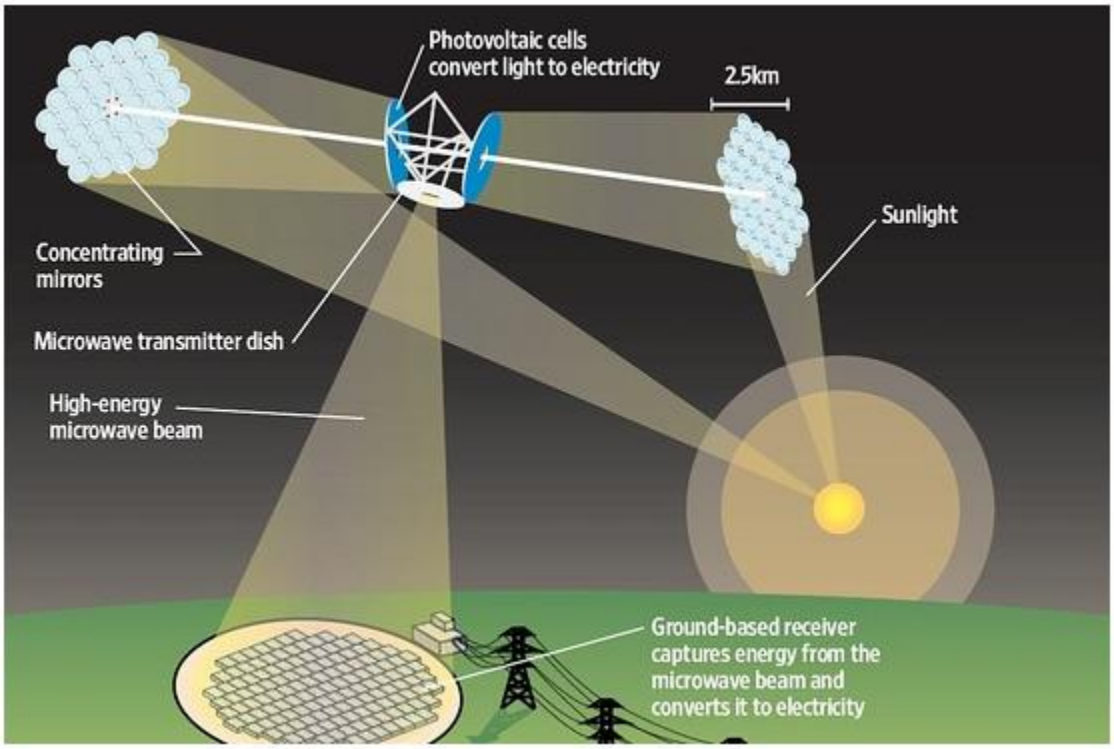


Fig. 2.1.2 Show the complete diagram of the SPS concept [2].

2.1.2.3 Progress Associated with Initial Interest in the Solar-Power Satellite (SPS) Concept

In 1968, Dr. Peter Glaser introduced the concept of SPS (Fig. 2.2). In this concept, the sun's energy is captured in orbit, converted into microwave power and then beamed to the Earth where subsequently it would be converted into electric power. This idea caught the attention of many companies interested in developing this project that seemed to be a possible solution for the growing demand for energy. A team of four companies (Arthur D. Little, Raytheon, Grumman Aerospace and Textron) was formed to develop a technical and economic feasibility study of the project. After six months of study, in 1971, the results of the technical investigation with favorable results encouraged them to send a letter to the director of NASA suggesting support for the SPS concept.

In 1973, the NASA's Office of Implementation, showed support for any additional development and also an special interest in the demonstration of the part of the system that

was responsible for power transmission. In 1975, the Goldstone demonstration was carried out, using a rectenna of 88 square meters, the power was transferred over a distance of 1 mile and a DC output of 30 kW was obtained; for the NASA and aerospace communities this demonstration was very important because it gave credibility to the power transmission and its use in the SPS concept. Concurrently, Raytheon developed a study sought to determine the critical technology areas of the SPS concept; this study concluded that these areas were: the DC to microwave conversion and the phased array antenna in the satellite. The development and improvement of these areas is considered the basis for many other technologies used in future applications.

A study supported by LERC in 1976-77, resulted in significant improvements in electrical and mechanical rectenna technology. The most important improvement was the mechanical change of format of rectenna, moving from a three-plane system, where bussing is done behind the reflector plane, making it more expensive, heavy and complex, to a system of two planes where there is only the reflector plane and a fore plane where all the microwave functions, rectification and DC bussing are held.

2.1.2.4 Later Period: Microwave Activity under DOE/NASA Concept Development and Evaluation Program for the SPS.

NASA, with the purpose of managing the energy that would be collected on Earth from the SPS system, involved ERDA (Energy Research and Development Agency) in the project. In 1977 ERDA began a three-year program called DOE/NASA Concept Development and Evaluation Program that resulted in a 670-page report summarizing the studies and findings. From these results, the most important for the development of microwave power transmission (MPT) was the suggestion of using the microwave oven magnetron. The program DOE/NASA SPS ended in 1980 with the conclusion that there was no reason for not continuing with the development and implementation of the project, but a program did not materialize. In retrospect, we can say that the greatest contribution of the SPS concept upon microwave power transmission (MPT) was the development of technologies that later replaced the classic super power tubes, for active phased array consisting of a large number of identical modules easy to manufacture using low-power microwave generators derived from the magnetron.

2.2 Low Power Energy Harvesting

Energy harvesting has been around in the form of windmills, watermills and passive solar power systems. In recent decades, technologies such as wind turbines, hydro-electric generators and solar panels have turned harvesting into a small but growing contributor to

the world's energy needs. This technology offers two significant advantages over battery-powered solutions: inexhaustible sources and little or no adverse environmental effects.

The most promising low power harvesting technologies extract energy from vibration, temperature differentials and light. Scavenging energy from RF emissions is interesting, but the energy availability is at least an order of magnitude less than the first three. Table 2.1 shows the estimated harvested power for different kinds of energy sources; we can observe that the estimated RF/EM harvested power along with the human vibration/motion harvested power, are the lowest values. Generally energy harvesting suffers from low, variable and unpredictable levels of available power.

The large reduction in power consumption achieved in electronics, along with the numbers of mobiles and other autonomous devices are continuously increasing the attractiveness of low power harvesting techniques. Consequently the amount of research in the field, and the number of publications has risen greatly.

Energy harvesting devices are particularly attractive as replacements for batteries in low-power wireless electronic devices. The cost of procuring, storing and getting someone to change a battery can easily cost as much as an energy harvester. A recent study [3] found that energy-harvesting systems are cost-comparable with systems using long-life batteries, over a ten-year lifetime.

2.3 EM/RF energy harvesting

When we refer to EM/RF harvesting, we do not refer to energy sources that have been specifically designed for powering wireless devices, we talk about the energy that we can collect from public services. In cities and very populated areas there is a large number of EM/RF sources like broadcasting radio and TV stations, mobile telephony base stations, and wireless networks. It is possible to collect part of this energy and convert it into useful energy. For our application, the main interest is in telecommunication services operating in the microwave region of the frequency spectrum, especially Global System for Mobile Communications (GSM) and Wireless Local Area Network (WLAN). As the most commonly used frequencies are well known, low-power devices have an antenna and a rectifier circuit (Rectenna), optimized to maximize energy harvesting at these frequencies[4]. For these services and their frequency bands, we can design printed antennas with dimension in the order of a few cm^2.

In laboratory environments, efficiencies above 90% have been observed, but when harvesting energy in the GSM or WLAN band on real scenarios, one has to deal with very low power density levels [5].

Table 2.1: Energy Harvesting Estimates [6] [7]

Energy Source Harvested Power	Harvested Power
Vibration/Motion	
Human	4 μW/cm^2
Industry	100 μW/ m^2
Temperature Difference	
Industry	1 – 10 mW/ cm^2
Light	
Indoor	10 μW/ cm^2
Outdoor	10 mW/ cm^2
RF/EM	
GSM	0.1 μW/ cm^2
Wi-Fi	0.001 mW/ cm^2
Solar	
Outdoor	10 mW/ cm^2
Indoor	< 0.1 mW/ cm^2
Acoustic	
75 – 10 dB of noise	0.003 – 0.96 μW/ cm^2
Human Body Sources	
Body heat	0.2 – 0.32 W (neck)
Walking	5 – 8.3 W

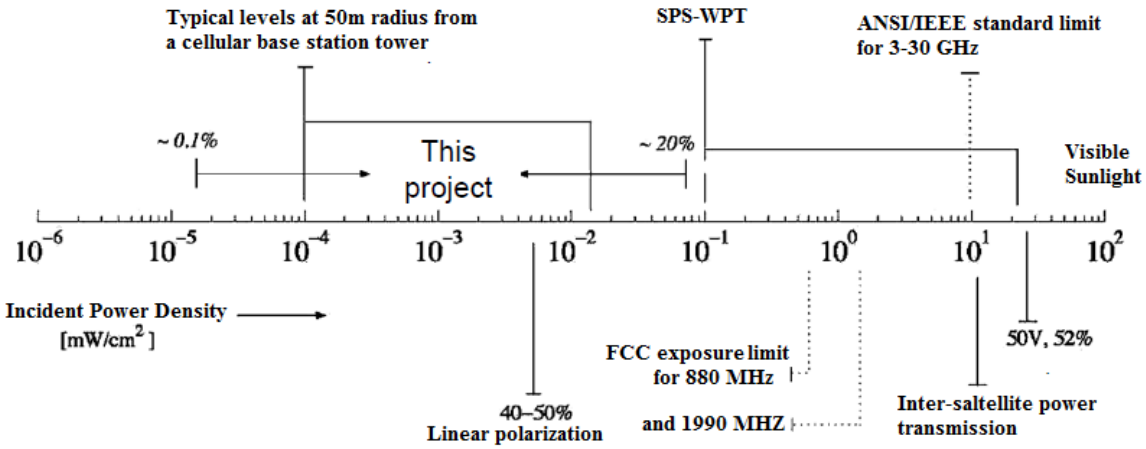


Fig. 2.2.1 Diagram of various microwave power sources and their typical power density levels. Also shown is the range of expected power densities used in the solar power satellite (SPS) and wireless power transmission (WPT) applications. The range of power densities for this work is indicated.

In Fig. 2.3.1 we see that the measured values of any kind of environment, at ground level, have a greater variation of power density, this is due to ground level, not always line of sight is obtained. For distances ranging from 25m to 100m from a GSM base station, power density levels ranging from 0.1mW/m2 to 1.0mW/m2 may be expected for single frequencies.

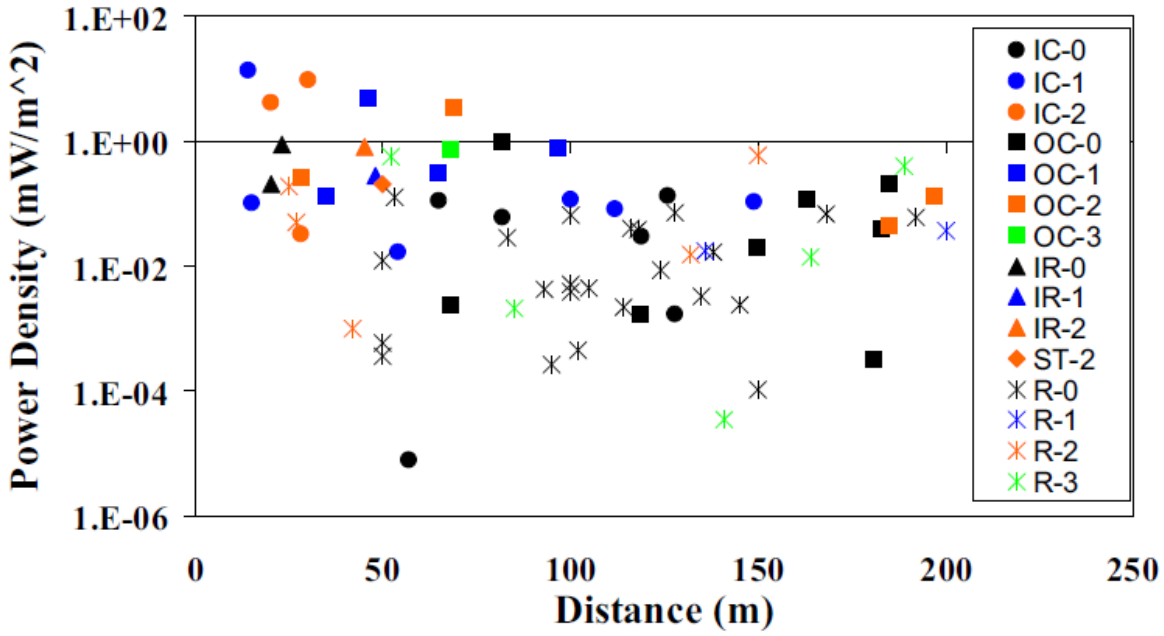


Fig. 2.3.1 Measured GSM-900 peak power density levels as a function of distance to the nearest base station. Data is taken from. Code "XY-a" indicates area and measurement site characteristics. XY: IC=Inner City, OC=Outer Country, IR=Industrial area, ST=Small Town, R=Rural or country-side are. "a": 0=outdoors on the ground, 1=outdoors on roof, terrace or balcony, 2=indoors, close to windows, 1.5m or less, 3=indoors, not close to windows.[8]

Figure 2.3.2 shows measurements over the entire 935-960 MHz band of the summed power density as a function of distance, we can conclude that between 25m and 100m, for indoor and outdoor environments, a high density power with values between 3.0 and mW/m2 0.3mW/m2 is expected. The measures over the entire frequency band are closely related to the traffic density at the moment of measurement, these measurements can vary by a factor form one to ten when compared with the measurements of a single Frequency (Fig. 2.3.1). Although the frequency of GSM-1800 band has doubled the GSM-900 band and therefore the free space loss has been quadrupled, ICNIRP (International Commission on Non-Ionizing Radiation Protection) exposures have only doubled. As for GSM-900, the actual exposures are well below the ICNIRP limits, we expect that power density levels for GSM-1800 are similar. This can be justified by measurements carried out in UK [9] and in Australia [10] for a mix of GSM-900/1800 base stations. From Figure 2.3.3, we can conclude that the power density measurements for GSM-900 and GSM-1800 are, up to 100 m, in the same order of magnitude.

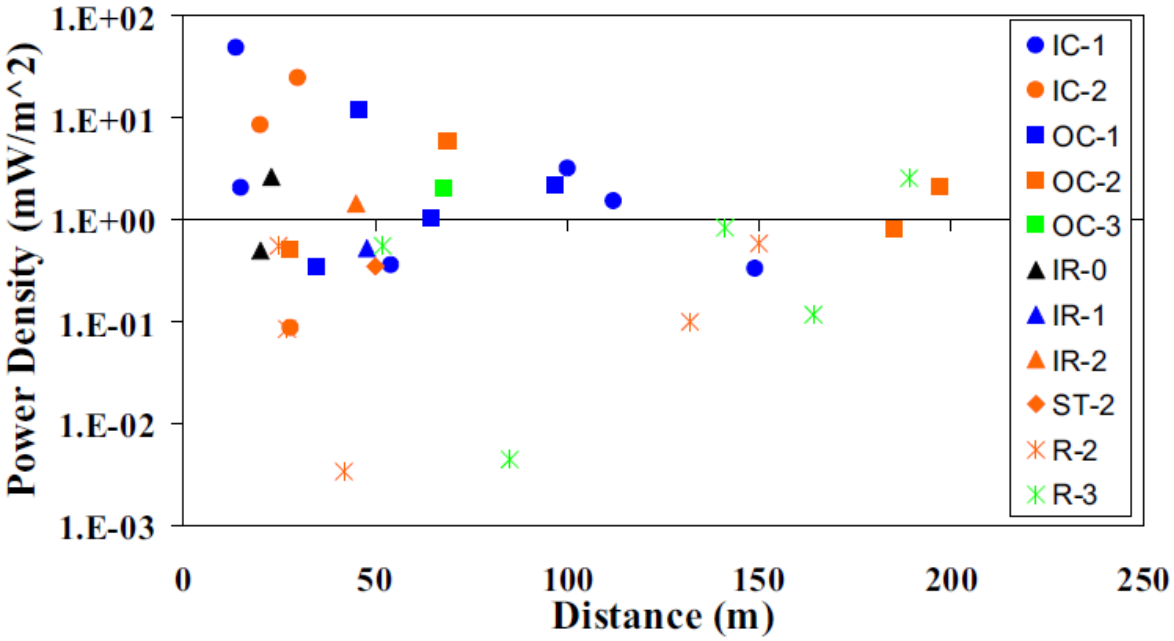


Fig. 2.3.2 Measured GSM-900 summed power density levels as a function of distance to the nearest base station, without data taken on ground level [8].

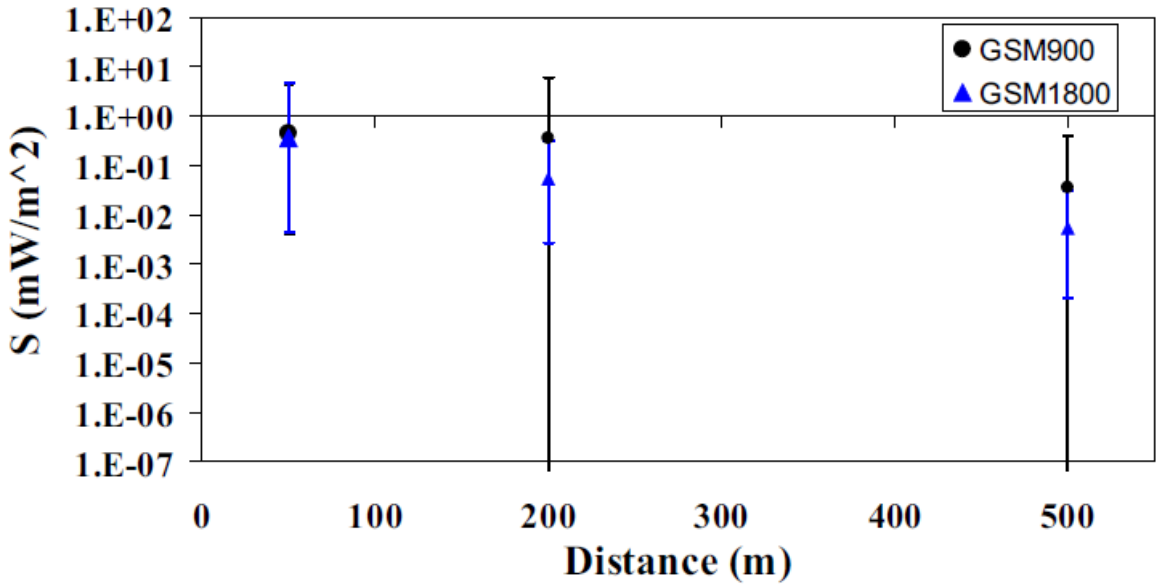


Fig. 2.3.3 Measured summed power density levels as a function of distance to the nearest GSM-900 and GSM-1800 base station [8].

If we refer to the 2.4 GHz band, whose efficiency in the environment is very high, Wi-Fi wireless local area networks (WLAN) applications are also useful with this concept. A WLAN router, transmit much less power than a GSM base station [8]. But because a router is confined to indoor environments and also that the distance would be much lower, reflections and low path loss may help to achieve suitable levels of power density. Fig.

2.3.4 shows the power density as a function of distance from the router. Although, during measurements of traffic levels were very low, we see that the power density levels are at least one order below the measurements obtained from GSM base stations.

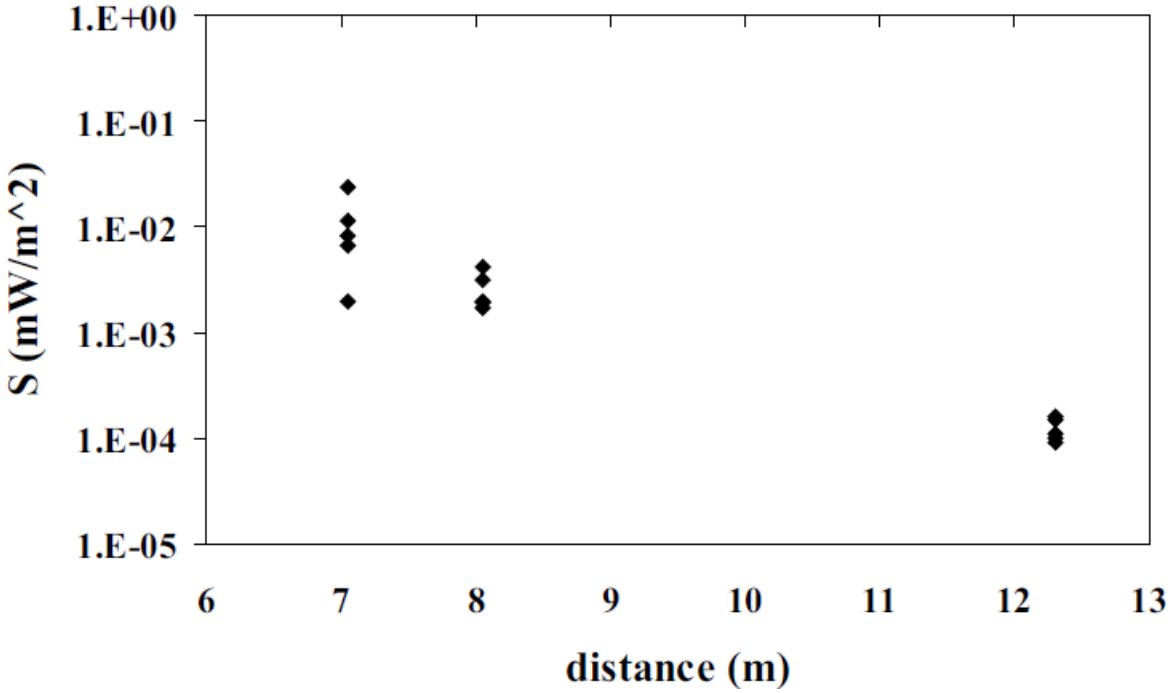


Fig. 2.3.4 Measured WLAN peak power density levels as a function of distance to the WLAN router [8].

2.4 Recent Technologies of Rectenna.

As explained above, a rectenna is a system that basically consists of a receiving antenna and a rectifier circuit which converts RF or microwave power into useful DC power. Fig. 2.1.2 shows the Schematic of a rectenna and associated power management circuit. The incident waves within a certain spectral range are received by the antenna, coupled to the rectifying device (diode in this case), and the low-pass filter (LPF) ensures that no RF is input to the power management circuit. A controller provides input to the power management circuit, which enables storage of the received energy over time, and delivery of DC power at the level and time when it is needed. This work will focus over the receiving antenna, the matching circuit and the rectifier circuit.

In recent years, a new approach to the use of rectennas has been reached and compared to other periods of microwave power transmission (MPT), where only high-power applications were developed (such as helicopter powering [1], solar-powered satellite-to-ground transmission [11], mechanical actuators for space-based telescopes [12], driving

motor [13], etc.), currently the develop of low-power applications is evolving. Rectennas implementing Ultra wideband (UWB) and narrow-band (essentially single frequency) antennas with smaller dimensions and new rectifier circuits configurations with new detectors that allow better performance are been used[14]. Self-powered sensors, short distances RFID tags, biomedical implants and energy harvesting devices, are some of the new low-power applications.

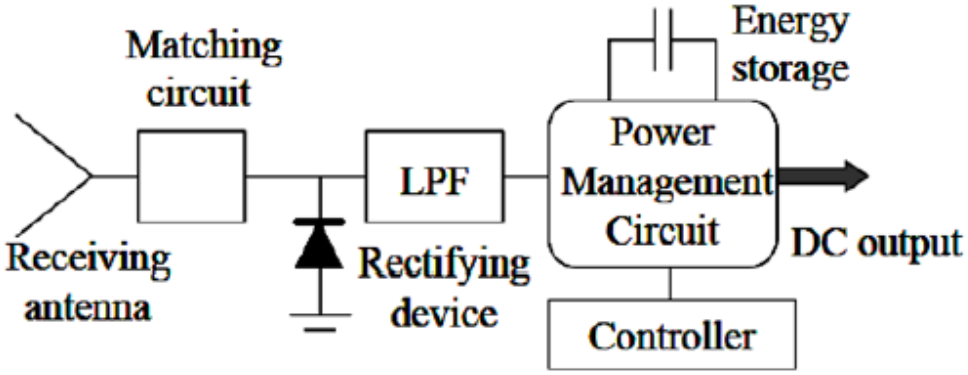


Fig. 2.4.1 Basic rectenna schematic with the power management system

Chapter 3

Antenna Design

In this chapter is described the design process, simulations and fabrication of an Ulta-WideBand antenna that operates from 900 MHz up to 6GHz. First we start with a brief explanation of the main design parameters of an antenna then we briefly explain the advantages of using UWB and microstrip antennas in these applications. Also a brief introduction of the software used for the design and simulations stages, as well as the method used for the simulator will be presented.

A parametric study of semicircular printed monopoles is included. Then we will explain how making some changes to the shape of the ground plane and the radiating patch of the antenna can improve the performance. After all the parametric simulations a final antenna version was chosen and subsequently built. Some pictures of the milling machine and the procedures after the milling are shown.

To improve the size of the antenna, a final version with some miniaturization techniques was built and the parametric simulations are included in a final study.

3.1 Antenna characteristics

To design an antenna, previously it is necessary to establish certain performance parameters that will be our goals in the design process and that are specific to our application.

3.1.1 Bandwidth

Bandwidth can be expressed as the range of frequencies over which the antenna characteristics are acceptable. In the case of broadband antennas, the bandwidth is often denoted as the ratio between the highest and lowest operating frequency. For example, 6:1 means that the higher frequency is six times greater than the lowest frequency.

3.1.2 Radiation Pattern

Is the mathematical representation or graphical representation of the radiation from an antenna as a function of spatial coordinates. Some of the properties of radiation are power flux density, radiation intensity, field strength, phase and polarization directivity, being the most important the spatial two and three dimensional radiation energy. A stroke of the power received at a constant radius is called power pattern. On the other hand, a plot of the spatial variation of an electric field (or magnetic) at a constant radius is called field amplitude pattern.

In practice, a three dimensional representation is obtained through various cuts of two-dimensional patterns. However for most practical applications, only some measures in terms of THETA for certain values of PHI and some measures in term of PHI for some values of THETA give us more useful information.

3.1.3 Radiation Power Density

Electromagnetic waves are used to carry information from one point to another through a wireless medium or through a guide structure, then it is natural to assume that the power and energy are associated with electromagnetic fields. The amount used to describe the power associated with an electromagnetic wave is the instantaneous Poynting vector

$$\mathcal{W} = \mathcal{E}\ x\ \mathcal{H}$$

$$\mathcal{W} = instantaneous\ Poynting\ vector\ \left(\frac{W}{m^2}\right)$$

$$\mathcal{E} = instantaneous\ Poynting\ vector\ \left(\frac{V}{m}\right)$$

$$\mathcal{H} = instantaneous\ magnetic\ field\ intensity\ \left(\frac{A}{m}\right)$$

Since the Poynting vector is a power density, the total power crossing a closed surface can be obtained by integrating the normal component of Poynting vector on this surface.

$$\wp = \oiint_S \mathcal{W}.ds = \oiint_S \mathcal{W}.\check{n}\, da$$

$$\wp = instantaneous\ total\ power\ (W)$$

$$\check{n} = unit\ vector\ normal\ to\ the\ surface$$

$$da = infinitesimal\ area\ of\ the\ colsed\ surface\ (m^2)$$

3.1.4 Radiation Intensity

The radiation density in a specific direction is defined as the power radiated by an antenna per unit solid angle, is a far-field parameter and is the product of multiplying the intensity of radiation by the square of the distance.

$$U = r^2.W_{rad}$$

$$U = radiation\ intensity\ (\frac{W}{unit\ solid\ angle})$$

$$W_{rad} = radiation\ density\ \left(\frac{W}{m^2}\right)$$

The radiation density in a certain direction is defined as the power radiated by an antenna per unit solid angle, is a far-field parameter and is the product of multiplying the intensity of radiation by the square of the distance.

$$P_{rad} = \oiint_\Omega U d\Omega = \int_0^{2\pi} \int_0^{2\pi} U.\sin\theta\, d\theta\, d\phi$$

$$d\Omega = element\ of\ solid\ angle = \sin\theta\, d\theta\, d\phi$$

In an isotropic antenna U is independent of the angles THETA and PHI.

$$P_{rad} = \oiint_\Omega U_0 d\Omega = U_0 \oiint_\Omega d\Omega = 4\pi U_0$$

$$U_0 = \frac{P_{rad}}{4\pi}$$

3.1.5 Directivity

In an isotropic antenna U the directivity is independent of the angles THETA and PHI. Directivity of an antenna is the ratio between the intensity of radiation in one direction from the antenna with the radiation intensity averaged over all directions. The average radiation level is equal to the total radiated power divided by 4PII. If the direction is not specified it is assumed the maximum intensity of radiation. In other words, the directivity of an antenna is not isotropic in a particular direction is equal to the ratio between the intensity of radiation with the radiation of an isotropic antenna.

$$D = \frac{U}{U_0} = \frac{4\pi U}{P_{rad}}$$

If the direction is not specified, it implies the direction of the maximum radiation intensity (maximum directivity) expressed as:

$$D_{max} = D_0 = \frac{U_{max}}{U_0} = \frac{4\pi U_{max}}{P_{rad}}$$

$$D = directivity\ (dimensionless)$$

$$D_0 = maximum\ directivity\ (dimensionless)$$

$$U = \text{radiation intensity (W/unit solid angle)}$$

$$U_{max} = \text{maximum radiation intensity (W/unit solid angle)}$$

$$U_0 = \text{radiation intensity of isotropic source (W/unit solid angle)}$$

$$P_{rad} = \text{total radiated power (W)}$$

3.1.6 Polarization

Polarization of an antenna can be defined as the polarization of the radiated fields measured or "seen" from far fields, some of the typical classifications are the linear polarization and circular (Right / Left Hand).

Because of the reciprocity theorem, antennas transmit and receive in the same way, then a vertically polarized antenna will communicate with another antenna vertically polarized and an horizontally polarized antenna will communicate with another horizontally polarized antenna, if, for example, a horizontally polarized antenna try to communicate with another whose polarization is vertical, there won't be communication at all.

To describe in general terms the loss of polarization we use the Polarization Loss Factor (PLF):

$$PLF = \cos\phi^2$$

The PFL describes the loss ratio in function of a rotation angle between two linearly polarized antennas.

A circular polarization is desired in many applications because there is no loss of polarization as a function of antenna rotation angles.

If, for example, a circularly polarized antenna receives a linearly polarized antenna signal(or vice versa), we must recall that the circular polarization is actually the product of two linearly polarized waves orthogonal to each shifted 90 degrees, then the receiving antenna will receive only the linear part which is half of the total power.

$$PLF(linear\ to\ circular) = 0.5 = -3dBm$$

3.1.7 Axial ratio

The axial ratio is a parameter used to quantify the polarization of an antenna and to denote deviations from a desired polarization and can be defines as the ratio of orthogonality of E fields components. When an antenna has a perfect circular polarization, the axial ratio is 1 (or 0 dB), because you have E-field components of the same magnitude, if it's an antenna with elliptical polarization, the axial ratio is greater than 1 and if it is a purely linear polarized antenna, the axial ratio tends to infinity because one of the components of E field is zero. The value of the axial ratio is degraded at a greater distance of the mainbeam.

3.1.8 Gain

The term gain describes how much power is transmitted in the direction of peak radiation to that of an isotropic source. The gain of an antenna is defined as the ratio of power delivered to the antenna and the power that is actually radiated. Ideally, the radiated power is equal to the received power, but in real specifications of antennas, due to various types of losses (conduction losses and dielectric losses) as well as mismatching losses, the ideal value is never obtained.

3.2 UWB Antennas

Broadband and UWB antennas date back to the earliest days of radio [15]. The wide scale commercial deployment of ultra wideband (UWB) systems has led to increased interest in UWB antenna design. The requirements imposed by UWB systems place stringent demands on UWB antennas. For example, the spectral and impedance matching properties of a UWB antenna influence on an overall UWB system design. A well-designed UWB antenna can contribute to system performance by notching out undesired frequencies.

UWB wireless technology is most thought for some application like:

- Very high-data-rate and short-range wireless communication systems
- Coding for security and low probability of interception
- Rejection of multipath effect
- Modern radar systems
- Energy harvesting
- Self powering devices

Printed antennas, commonly fabricated on FR4 substrate, are very cost effective, which is ideally suited for UWB technology-based low-cost systems.

3.2.1 Microstrip Antennas

Microstrip antennas date back to the 50's when they were patented, but it is in the 70's when they receive considerable attention [16]. With the development of the high data rate wireless communication systems, antenna engineers are faced with increasing demand for smaller antennas which still retain broadband characteristics.

A microstrip antenna is formed basically by a dielectric substrate between two metal layers, one layer is the antenna and the other is the ground, Fig. 3.2.1.1 shows the basic configuration of a microstrip antenna

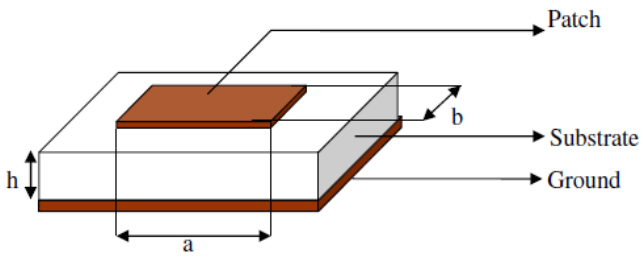


Fig. 3.2.1.1: basic layers of a microstrip antenna

Microstrip antennas, also known as patch antennas, performance is affected by the patch geometry, substrate properties and feed techniques, they often have the radiating elements and the feeding line etched in the dielectric substrate, they have its maximum radiation in the normal direction of the plane of the antenna that is achieved with an appropriate excitation below the antenna.

There are many types of substrate with dielectric constant values usually in the range of 2.2 <ER <12. Thick substrates with low dielectric constants are desired because they have higher bandwidth efficiency at the cost of a larger antenna size. On the other hand thin substrates with a high dielectric constant are desired in applications of microwave circuitry because they minimize unwanted coupling and radiation, and also because they allow to design smaller antennas. Since the microstrip antennas are closely linked to microwave circuits, a compromise between efficiency, bandwidth and size is necessary [16].

In recent years, printed planar monopole antennas have become very popular [16] and some of the main advantages of these antennas are:

- Low profile
- Low cost
- Light weight
- Ease of fabrication and installation
- Large bandwidth
- Radiation properties, wide impedance bandwidth and nearly omnidirectional azimuthal radiation pattern)
- Some of the various applications are covered with fewer or with a single antenna.

3.2.2 Feed mechanisms

Most of the performance and complexity of the design of a microstrip antenna depends on how the antenna patch will be fed. We can classify the techniques of feeding in three large groups.

- *Directly connected*: in this technique, a connection is made directly to the radiator element. For example, a microstrip line or a coaxial cable. This technique is very efficient in thin undertreated and one advantage is that the feeding line and the patch antenna are on the same plane, so it becomes a planar construction. In the case of a coaxial cable, it connects to the patch antenna through a hole in the substrate and the outer shell is grounded by merging it with the ground plane of the antenna.
- *Aperture Coupling*: This technique was introduced to overcome the technical problems of the directly connected feeding technique. It basically consists of two

substrates separated by a ground plane; in the bottom of the lower substrate, there is a microstrip line whose energy is coupled to the patch that is in the top of the upper substrate through a hole in ground plane. This feeding is the most difficult to manufacture and has a narrow bandwidth, but it is easier to model and has moderate spurious radiation [16].

- *Proximity Coupling*: The technique of proximity coupling consists of a power line located between two substrates, at the bottom of the lower substrate is the ground plane, and at the top of the upper substrate is the patch. Of the four feeding techniques, proximity coupling is the one with the higher bandwidth, is relatively simple to model and the spurious radiations are low, however its manufacturing is more difficult.

3.2.3 Software

For the stage of physical design and simulation, the Ansoft HFSS 11 software has been used. In order to generate an electromagnetic field solution, HFSS employs the finite element method. Fig. 3.2.1.1 shows the main window of the HFSS11; in the right side, a preview of the design is shown, the middle part, lists the components that make up our design, such as, solids, sheets, and planes. The left side shows the configurable features of the project such as the frequency sweep settings and simulation results.

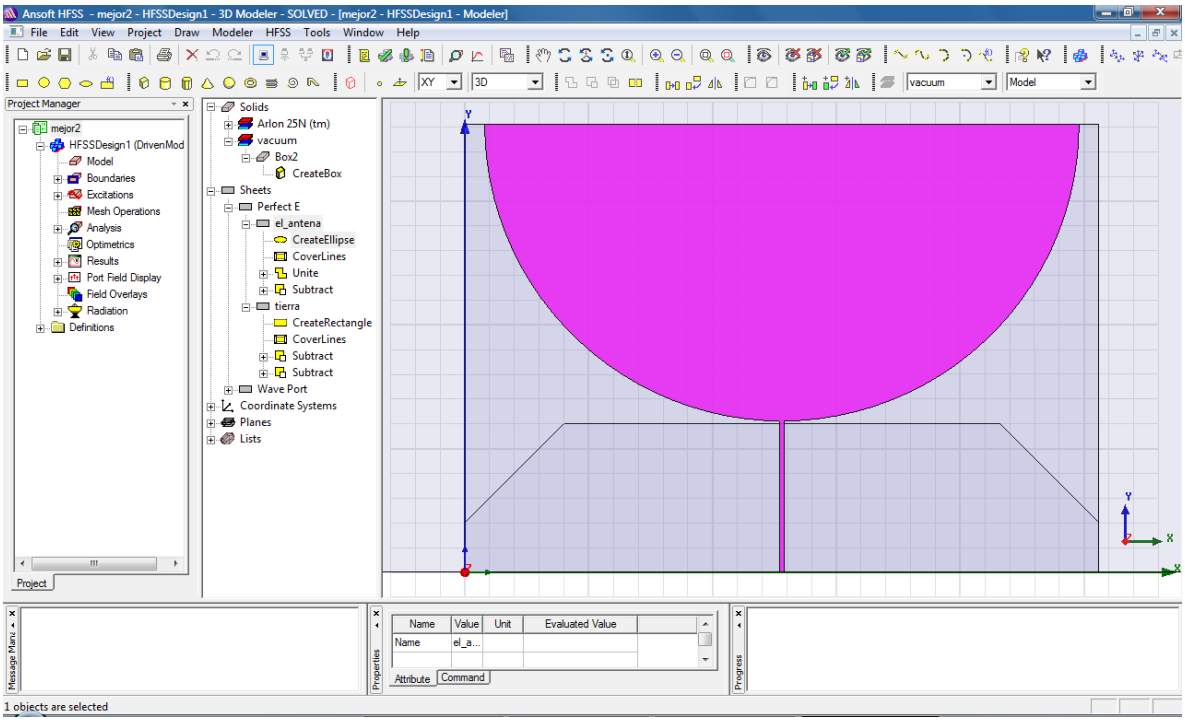


Fig. 3.2.1.1: Main HFSS11 window.

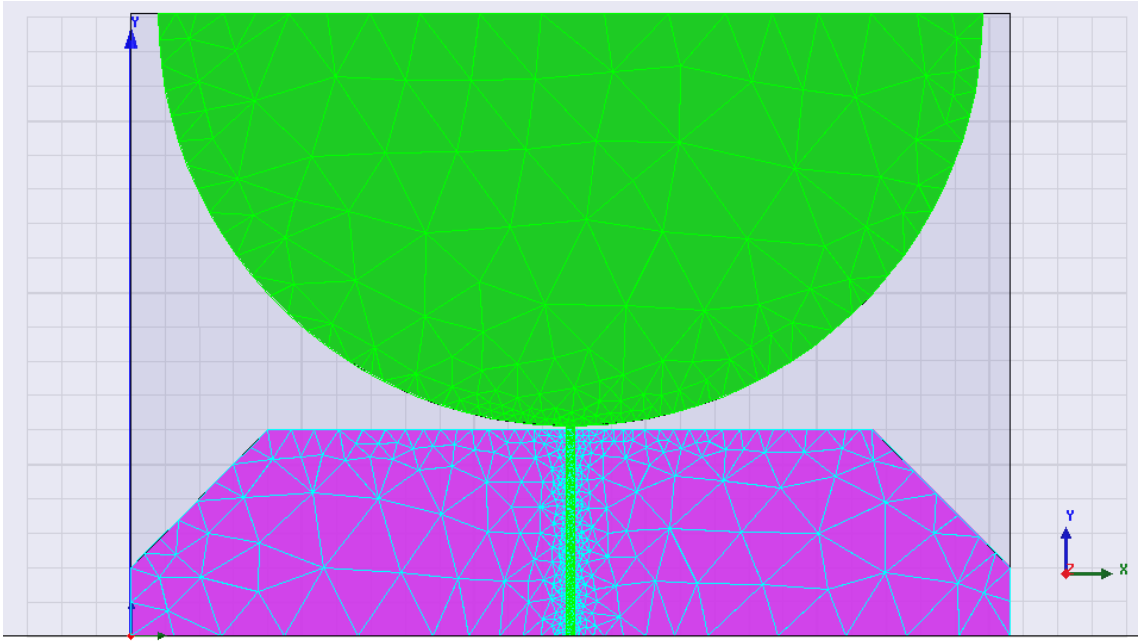


Figure 3.2.1.2: Generated mesh to solve the antenna design employing the finite element method

In general, the finite element method divides the full problem space into thousands of smaller regions and represents the field in each sub-region (element) with a local function. One feature of this method is that in areas of interest, the resolution of elements (Mesh) becomes denser in order to obtain more accurate predictions. Fig. 3.2.1.2 shows an example of how the antenna is divided into smaller sections and then apply the finite element method and Fig. 3.2.1.3 shows the higher density of elements.

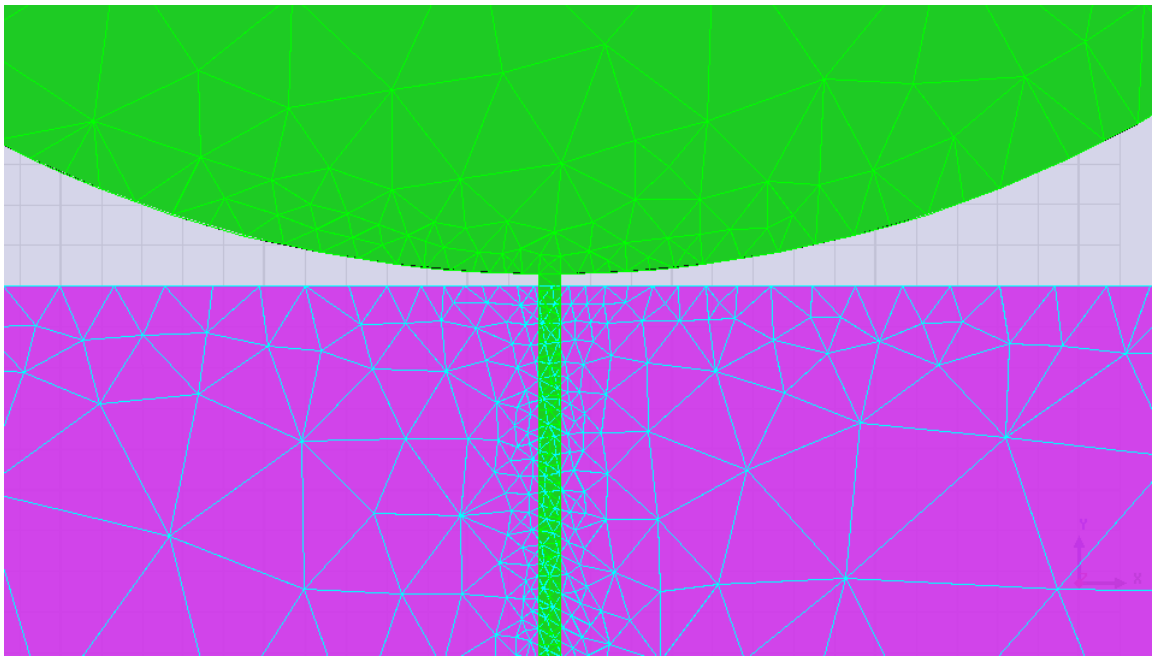


Figure 3.2.1.3: Higher density of elements to obtain a more accurate result

3.3 Antenna Simulations

In this section, we will explain the process for the design and choice of two UWB antennas that were subsequently built. A bandwidth covering form 900 MHz - 6 GHz (so that the main 900/1800 GSM downlink bands and the 2.4 - 2.5 GHz ISM band, were covered) and not very large physical dimensions were the main characteristics desired for our antennas. The results of the simulations and parametric studies will be shown and analyzed. According to these studies, the best antenna was chosen and then, a modification to the ground plane was made; this analysis will be detailed later. After this last modification, the antenna was built and we will call it, from now on, base-antenna.

In order to reduce the size of the base-antenna, but maintaining the bandwidth, miniaturization techniques were applied, a parametric study of this process will be included and explained, to be then built; we will call this antenna, from now on, size-improved antenna.

For the design of our antennas, we have started with a single shape, and, according to the simulation results and references, we have tuned some parameters to reach the best possible performance. The table 3.3.1 shows the parameters of the available substrate.

Table 3.3.1: Dielectric parameters

	Arlon 25N (A25N)
Er	3.38
TanD	0.0025
H	0.508mm
Cond (copper)	5.813×10^7

3.3.1 Base Antenna

For antenna design we have not followed a parametric design procedure. The design has started with a pre-existing half disc antenna and we have adapted it to our needs through simulations. We have varied the parameters shown in Fig. 3.3.2.1. In general terms, the main differences between a full disc antenna and the half disc are that low frequency limit of the band increases slightly, but the phase response remains linear, also, the received peak amplitude is also deceased by about 20% from the disc antenna [18]. This is not a large penalty for obtaining the substantial shortening of the height to half of the original disc antenna. A small size and a broadband antenna are our main goals.

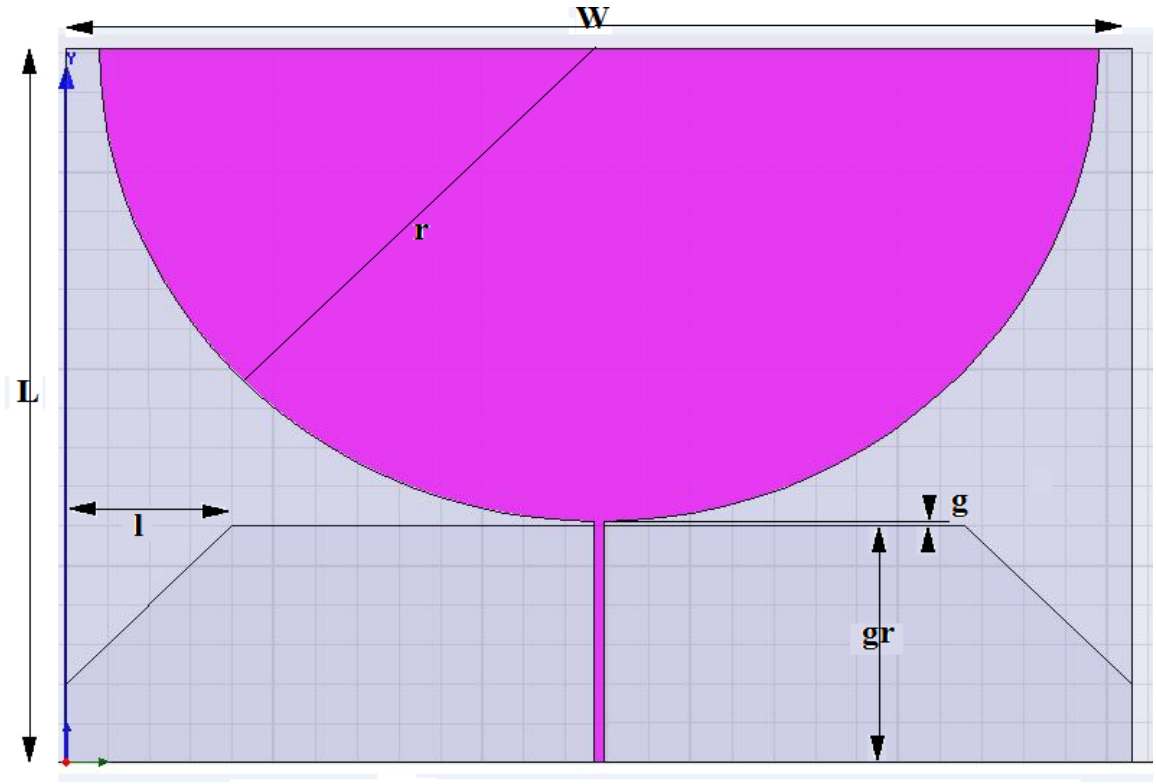


Fig. 3.3.1.1: 2D view of the base-antenna.

As if it were a circular monopole patch, the radius r of the circle is the main parameter to optimize for the lowest return loss and widest bandwidth. We will do a parametric sweep of "r" while the other parameters are kept constant. Fig. 3.3.1.1 shows the behavior of changing this parameter; we can observe that as the radius of the circle increases, the lower frequency decreases.

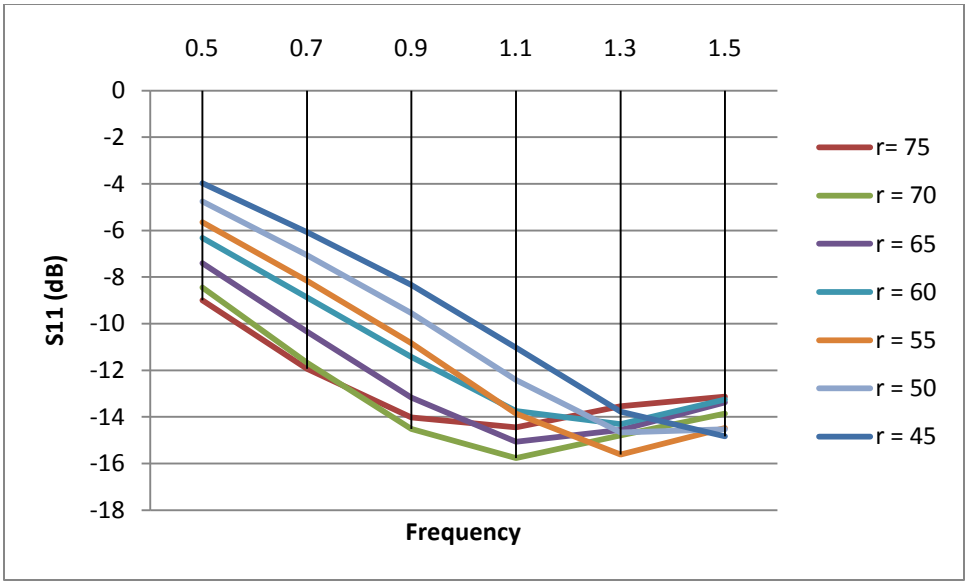


Fig. 3.3.1.2: low-frequency return-losses as a function of the "a" parameter

The following parameter to tune is the gap between the patch and ground plane ("g"). Parameter is stated to:

- Have positive value if the patch is at a higher level than the ground layer.
- Have null value when the ground plane ends where the patch begins.
- Have a negative value if the patch and ground layer overlap.

Since the simulation software does not accept negative values, we have denoted the parameter "h" as "g + gr", and the ground plane height is 30mm.

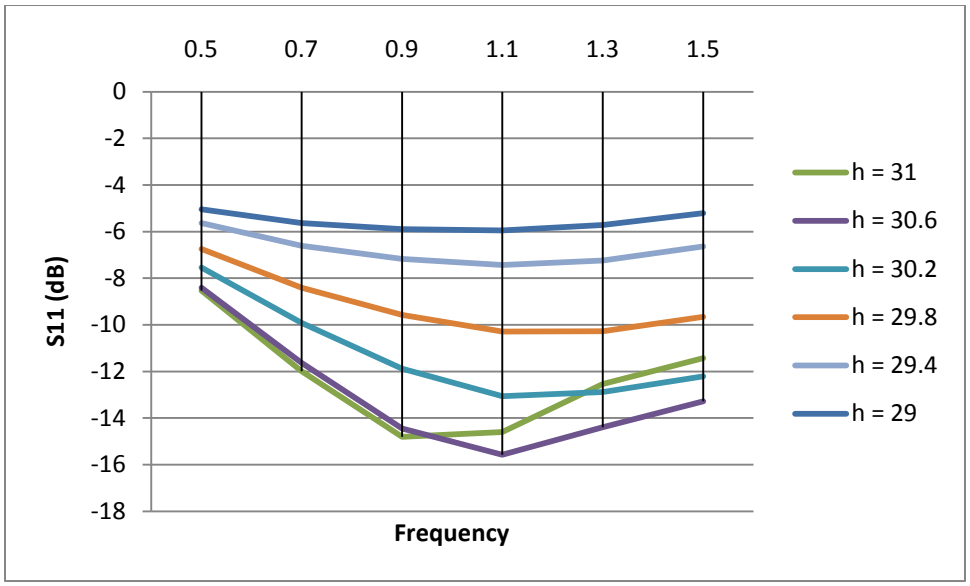


Fig. 3.3.1.3: low-frequency return-losses as a function of the "a" parameter

From Fig. 3.3.1.3 we can conclude that the best performance value is h = 30.6 mm (g = 0.6 mm)

3.3.2 Ground Modification

To improve the bandwidth of the antenna, we have removed the upper corners of the ground plane and we will vary them to reach the best performance. Fig. 3.3.2.1 shows the results of the variation of the "l" parameter. It also shows that "l" has not considerable effect at low frequencies. But at frequencies above 1 GHz is a noticeable difference.

Fig.3.2.2.3 shows the radiation patters for each frequency and Fig. 3.2.2.4 represents the gain pattern of the base-antenna

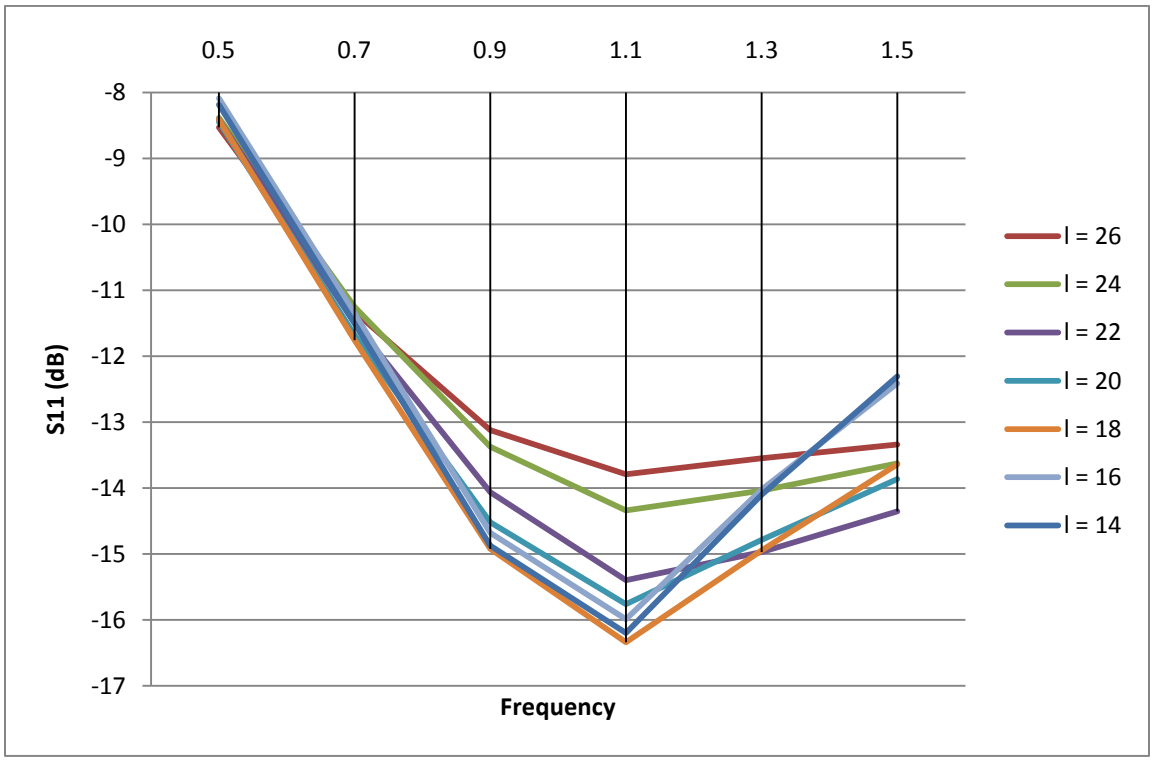


Fig. 3.3.2.1: low-frequency return-losses as a function of the "l" parameter

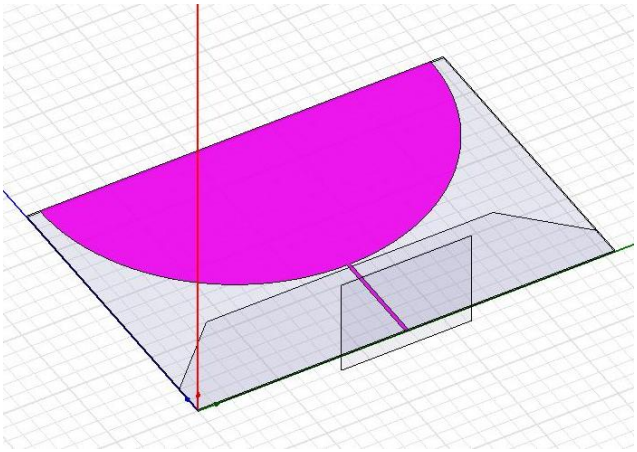
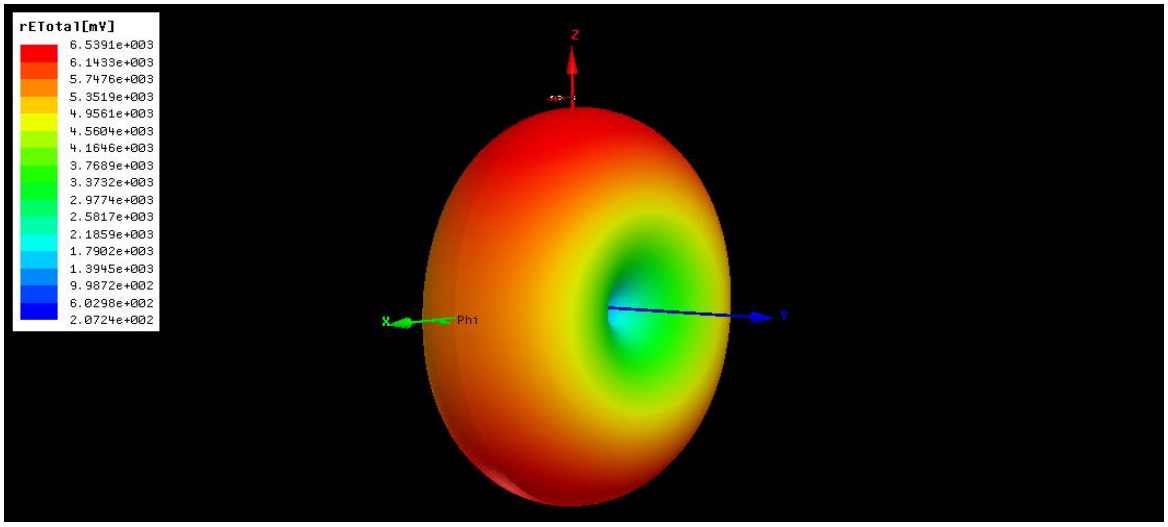


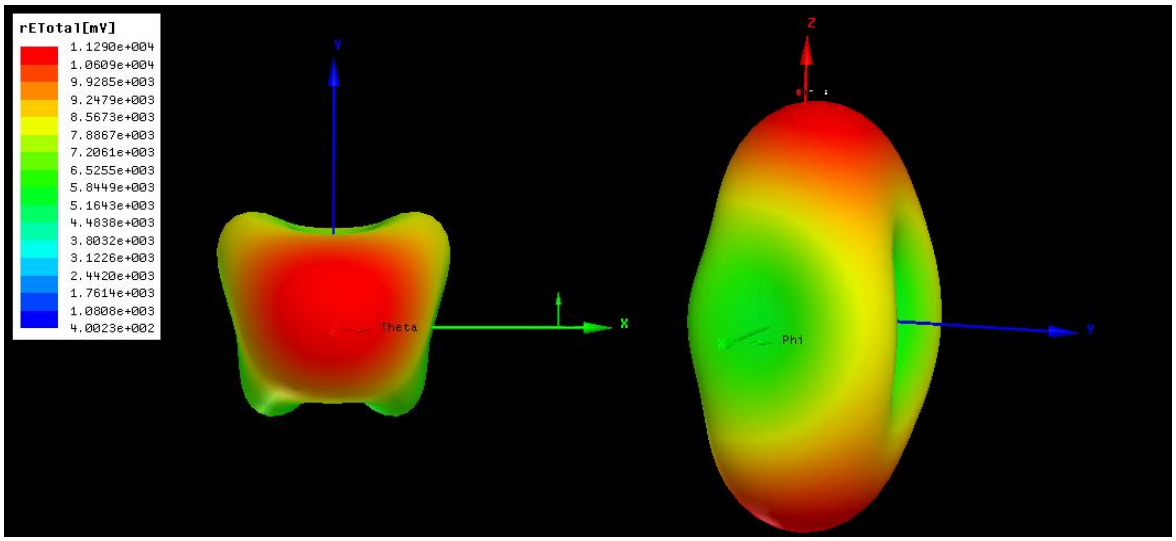
Fig. 3.3.2.2: 3D view of the base-antenna (HFSS11)

Table 3.3.2.1: Final values for the base-antenna, the name of the values are taken from Fig. 3.3.2.2.

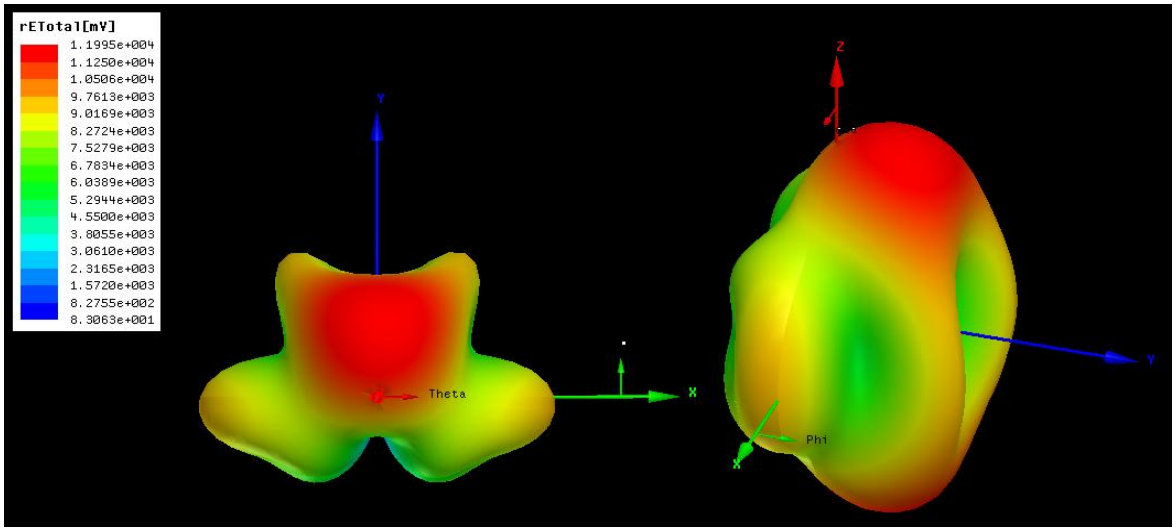
Name	Value
L	100,6 mm
W	148 mm
l	20 mm
gr	30 mm
g	0.6 mm
r	70 mm



(a)



(b)



(c)

Fig. 3.3.2.3: Base-antenna radiation patterns at (a) 900 MHz, (b) 1.85 GHz and (c) 2.45 GHz

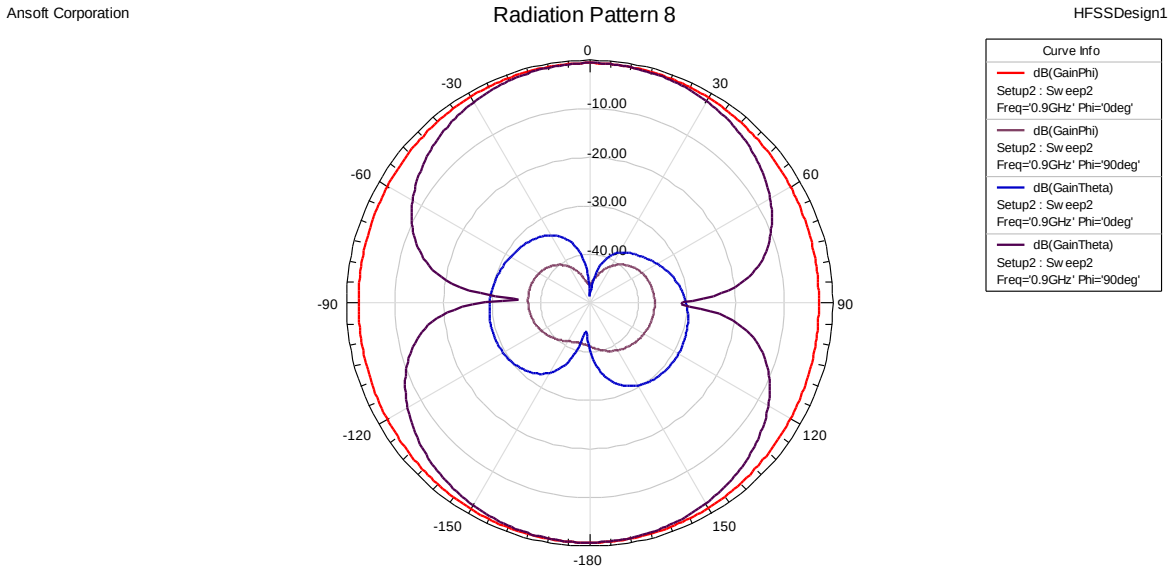


Fig. 3.3.2.4: base-antenna gain pattern at 900 MHz

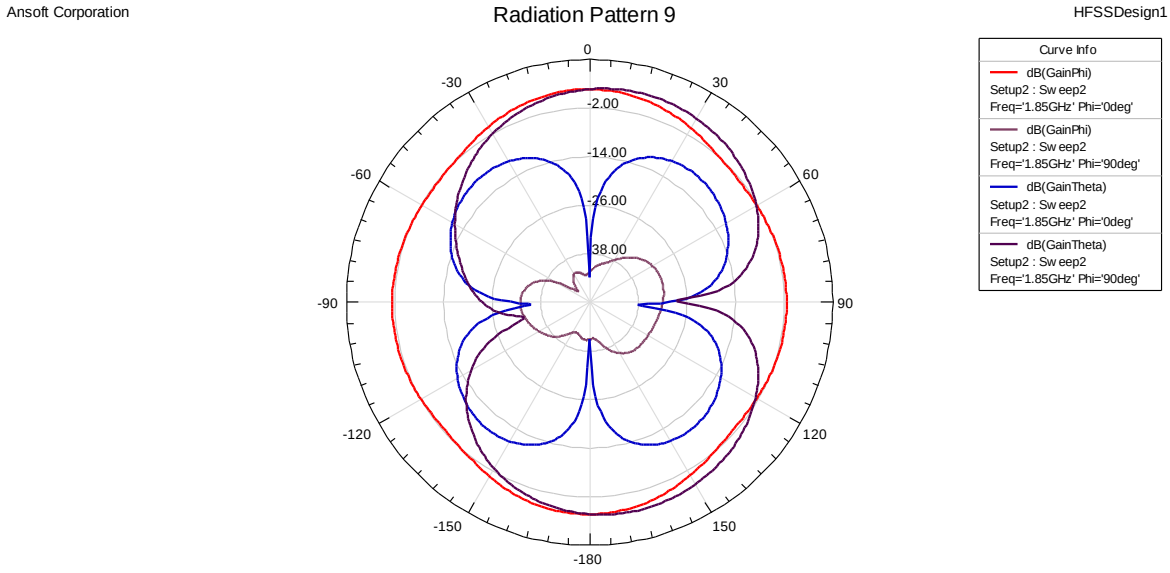


Fig. 3.3.2.5: base-antenna gain pattern at 1.85 GHz

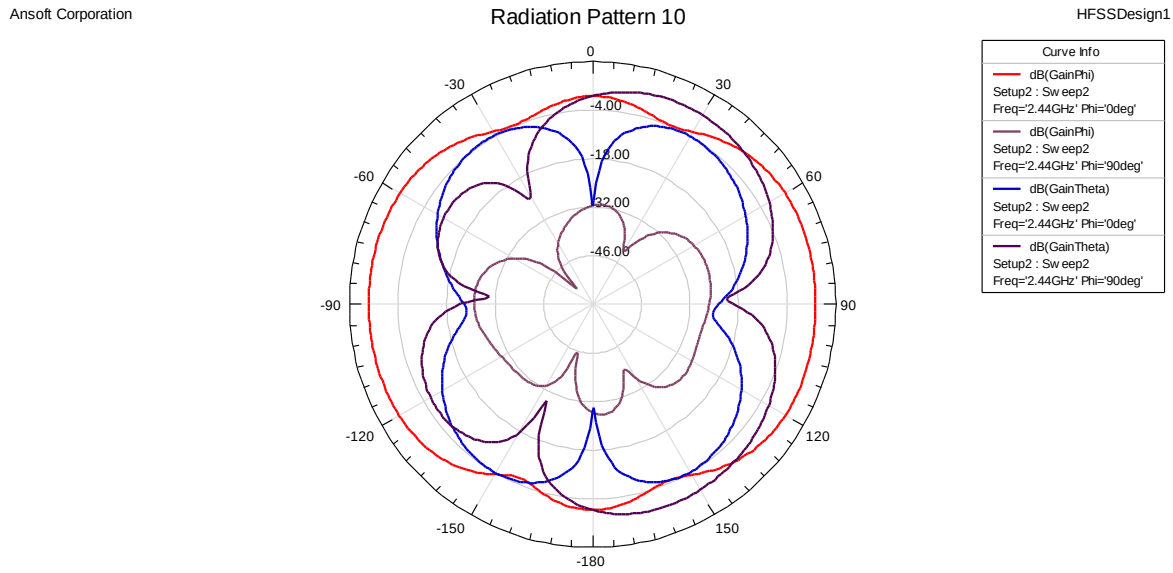


Fig. 3.3.2.6: base-antenna gain pattern at 2.45 GHz

3.3.3 Miniaturization Techniques

This section is intended to reduce the size of the design that we had previously. In [19], a series of procedures that aim to alter the current distribution at the antenna surface by inserting a slot is shown. As we have a different antenna (in the paper is a rectangular antenna), we cannot guide from the slot dimensions that are presented in this paper.

To overcome this, we'll perform simulations by varying some parameters to finally choose appropriate slot values.

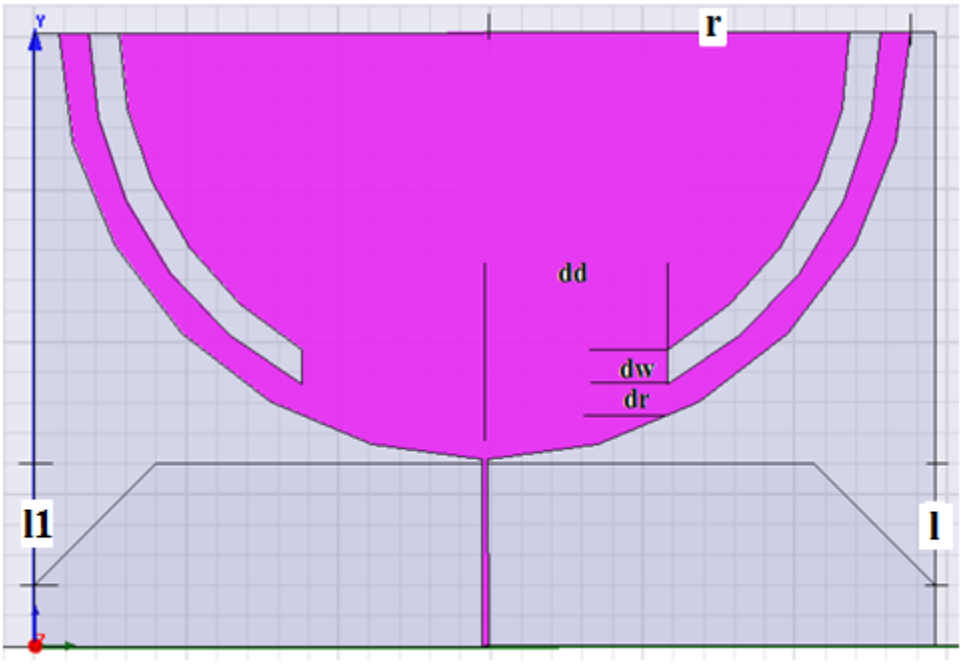


Fig. 3.3.3.1: Slotted antenna and its parameters

Fig. 3.3.3.1 shows the first slotted antenna prototype. The dimensions of the slot are denoted with the "dd", "dr", "dw" variables. We will vary each one of these parameters in a given range to see how they affect the antenna performance.

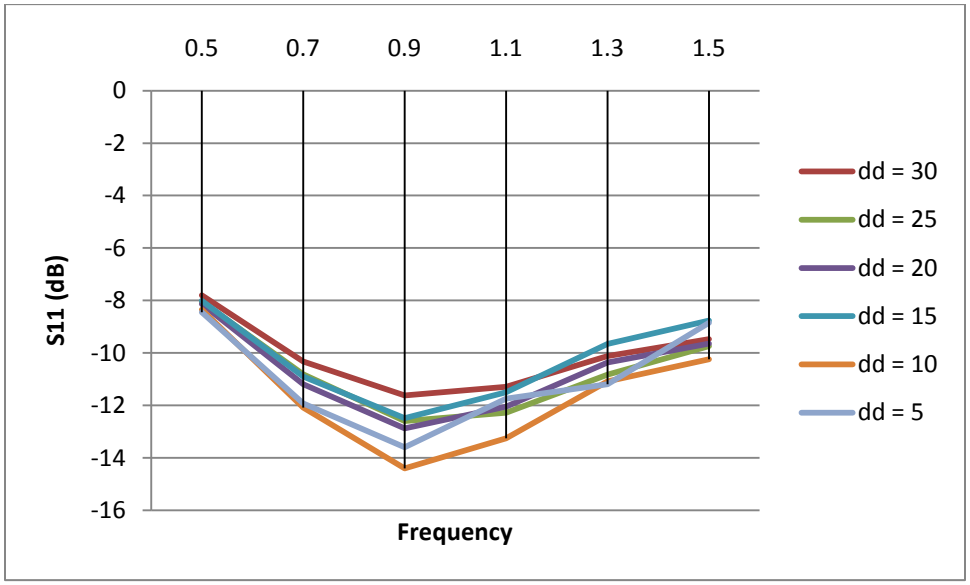


Fig. 3.3.3.2: low-frequency return-losses as a function of the “dd” parameter

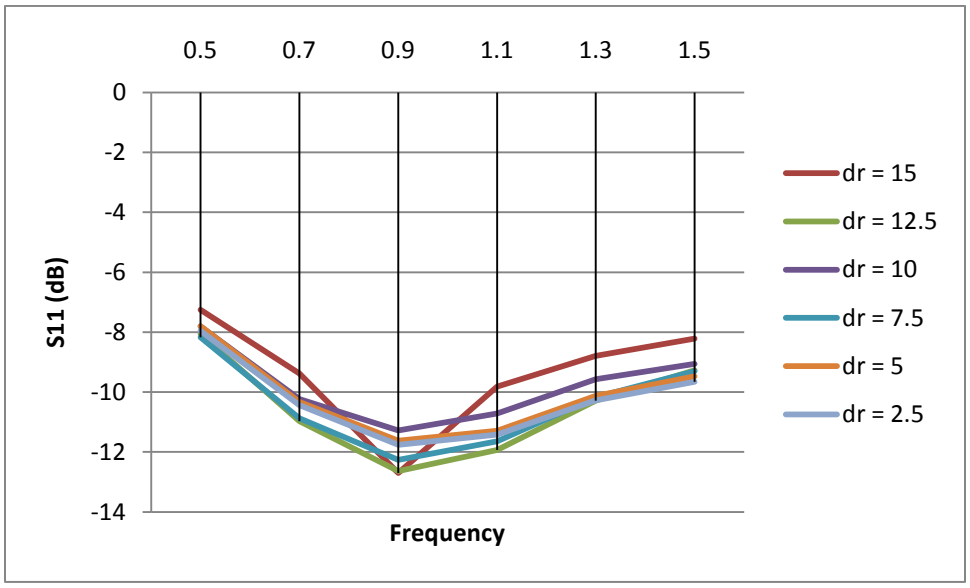


Fig. 3.3.3.3: low-frequency return-losses as a function of the "dr" parameter

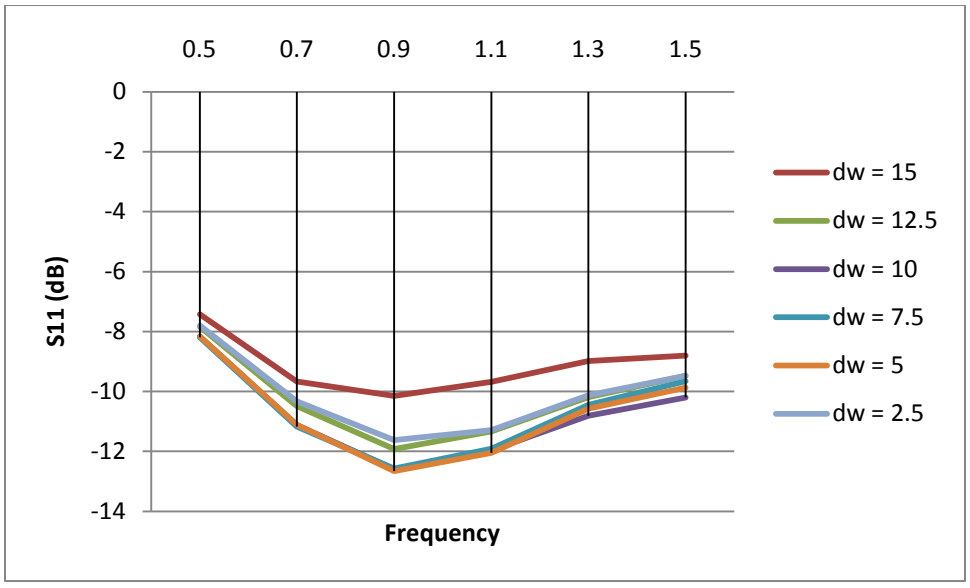


Fig. 3.3.3.4: low-frequency return-losses as a function of the "dw" parameter

After seeing these parametric simulations, we can observe that the slot length has effect on the impedance matching, especially at low frequencies. The width and location of the slot slightly influence the low frequencies. In general, all parameters related to the slot have some influence in the impedance matching.

The next step is to reduce overall the size of the antenna; in this step reduction will be done by reducing the radius "r" of the semi-circle. Because of the total dimensions of the substrate are as a function of the radius, they are automatically accommodated after a radius change. Fig. 3.3.3.5 shows how return loss varies by changing this parameter.

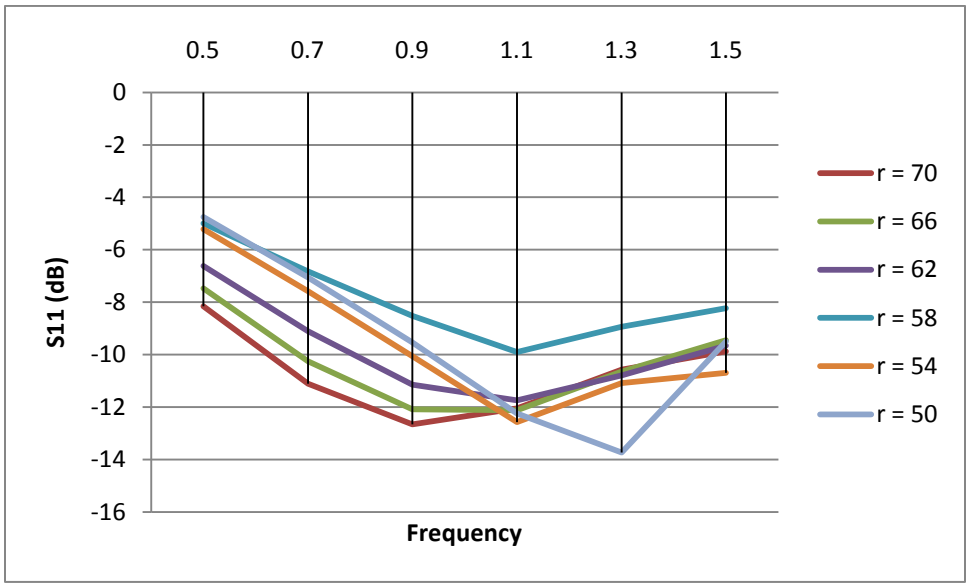


Fig. 3.3.3.5: low-frequency return-losses as a function of the "a" parameter

After seeing the results in Fig. 3.3.3.5, we have decided to reduce the radius "r" from 70mm to 60mm, with this adjustment, dimensions of the substrate, have been reduced from 148mm x 100mm to 120mm x 90.6, 19% and 9.4% respectively. Because of the change of th antenna, the response is not the same, but as we know the behavior of each parameter we have adjusted them and the final values now are: dd:14mm, dr:10mm and dw:7mm.

The next antenna design will intend to reduce even more the previous antenna design by inserting an angle cut as we know that current distribution is supposed to concentrate close to the slot (we will verify this in section 3.3.4), then, we will try to concentrate it in one slot. The next step is to make an angle cut to the patch. Fig. 3.3.3.6 shows the way we will do this, we will use the middle line of our antenna as a reference, and moving to the left, we will find the proper angle.

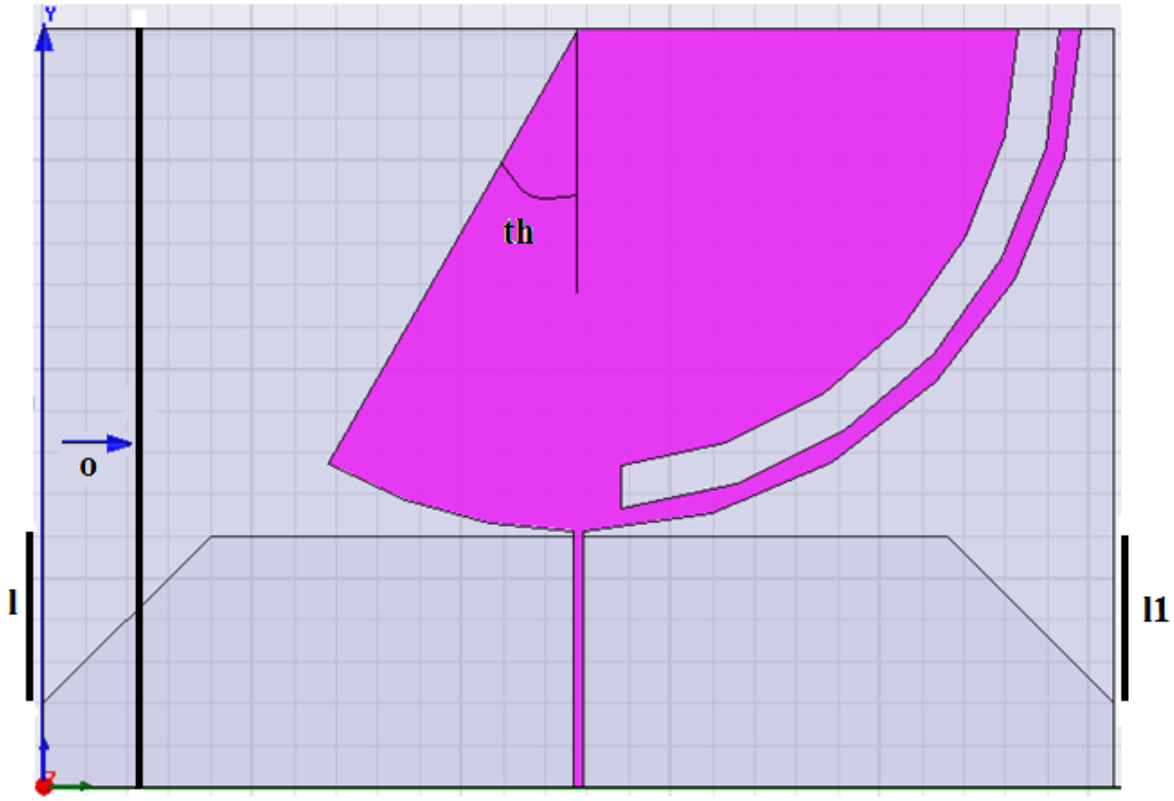


Fig. 3.3.3.6: shows the antenna and the "th" parameter

From Fig. 3.3.3.7, we can conclude that as the angle increases, frequencies above 1GHz change their performance dramatically. Lower values of "th" are desired, because they would provide us more free-area in the left side of the antenna.

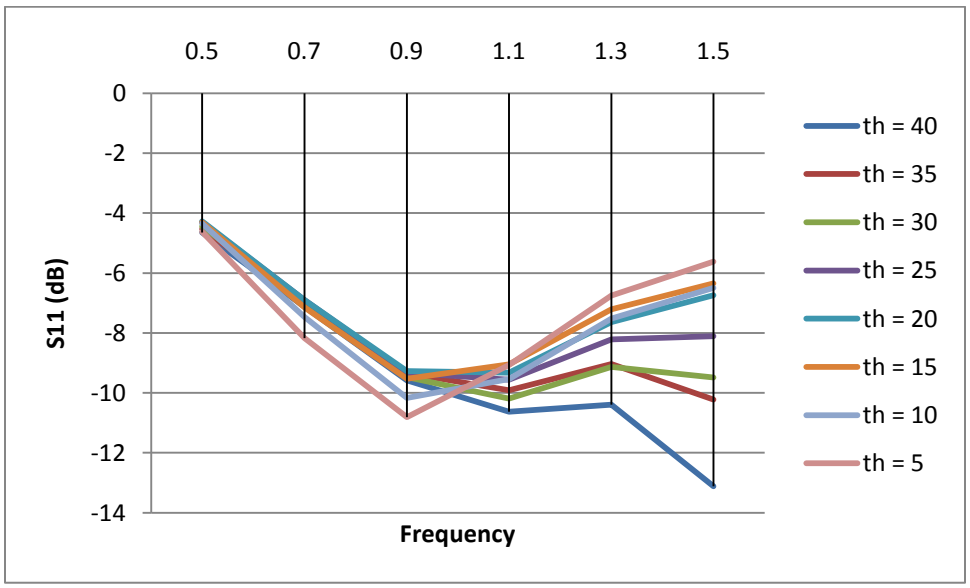


Fig. 3.3.3.7: low-frequency return-losses as a function of the "th" parameter

After choosing a particular angle of cut, the next step is to cut the substrate starting from the left edge. The parameter "o" indicates the number of millimeters that we will cut the substrate. For this simulation, we have determined that the angle "th" will be 30 degrees, as it is a good balance between return loss and space saving. Fig. 3.3.3.8 shows the response of the parameter "o", we can conclude that a nice cut value is 8mm.

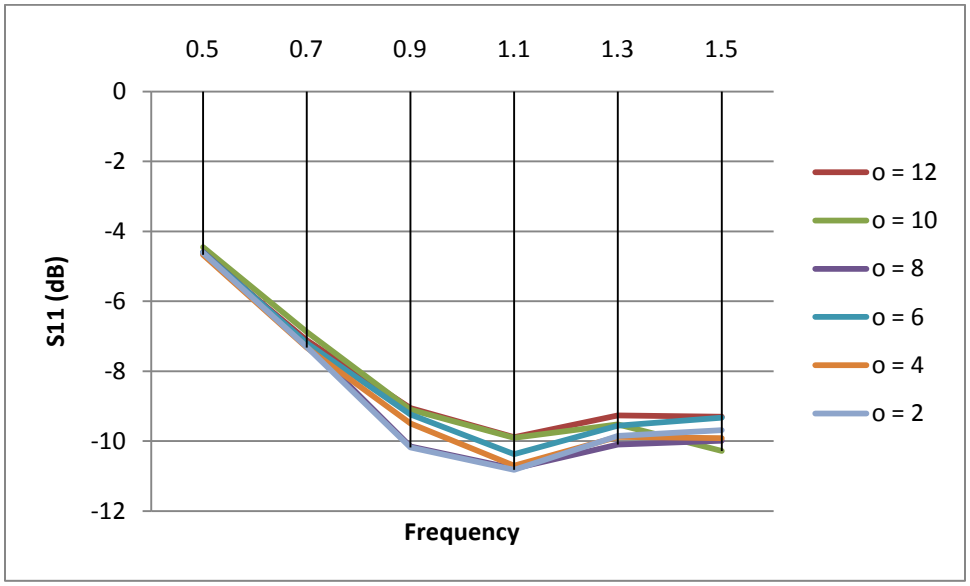


Fig. 3.3.3.8: low-frequency return-losses as a function of the “o” parameter

The next step is to try to improve the ground plane, for this issue we have tried the following (remember that until now, we have two edges, each one of 20mm at both sides of the antenna):

- Keeping the right edge (l = 20mm) and varying the value of the edge of the left (l1).
- Keeping the left edge (l1 = 20mm) and varying the value of the edge of the right (l).

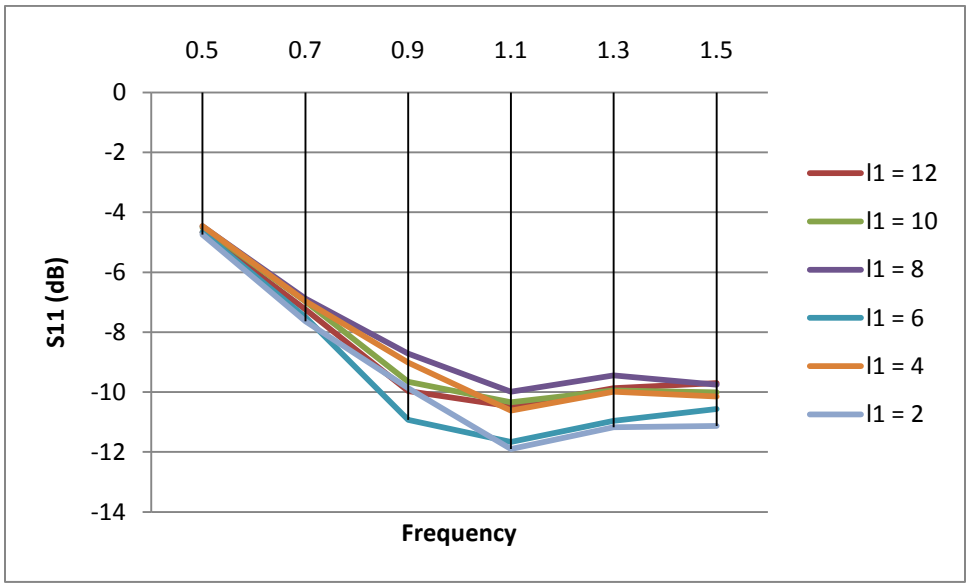


Fig. 3.3.3.9: low-frequency return-losses as a function of the "l1" parameter

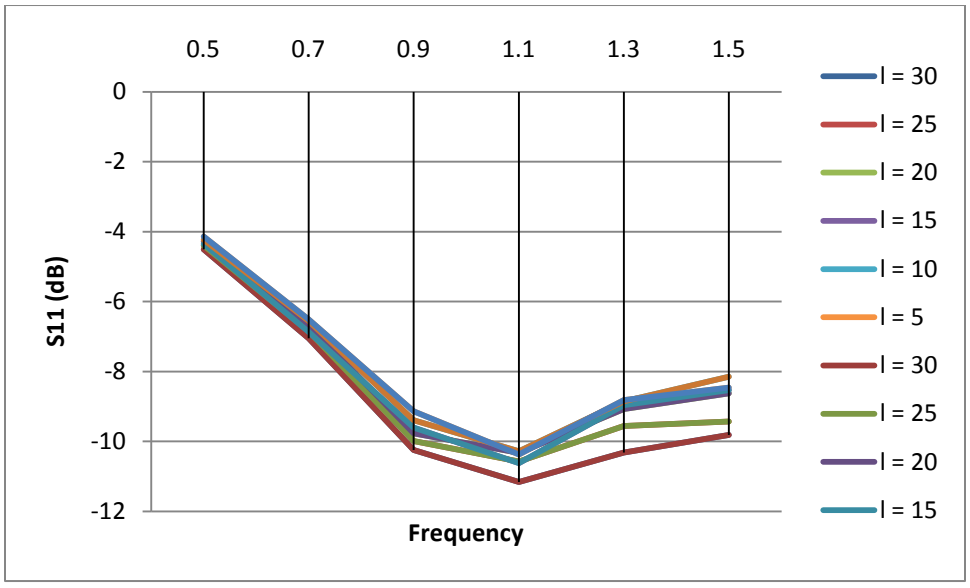


Fig. 3.3.3.10: low-frequency return-losses as a function of the "l" parameter

Because a variation of the left edge has provided better results, and moreover, a low value for the left side has shown a good performance (Fig. 3.3.3.9), we have decided to cancel the left edge and perform another simulation varying the right edge.

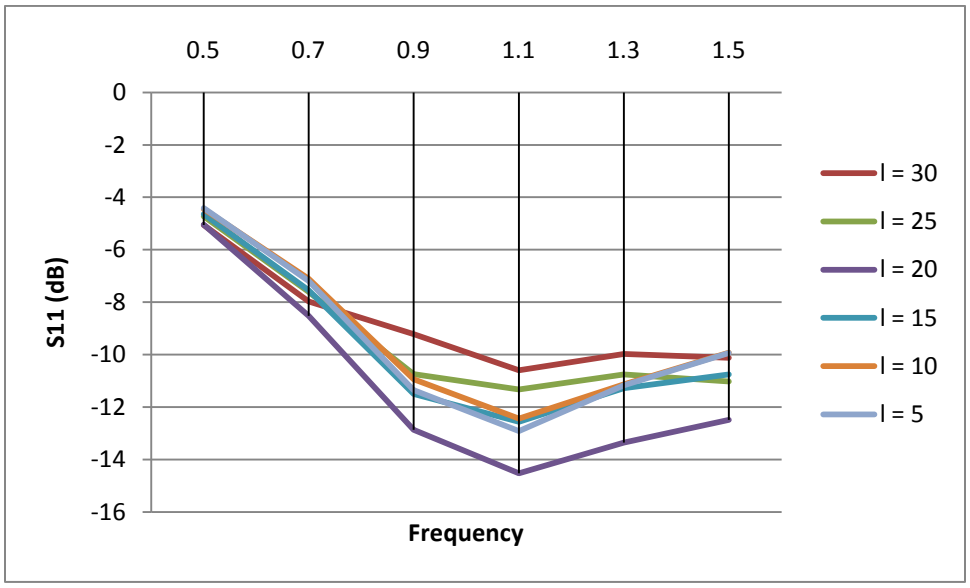


Fig. 3.3.3.11: low-frequency return-losses as a function of the "l" parameter

From Fig. 3.3.3.11 we can conclude that the best performance was obtained by removing the left edge and by maintaining the right edge with 20 mm side long.

3.3.4 Size-Improved Antenna

After having analyzed each of the previous settings and after making many simulations, we have concluded that the antenna to be manufactured would be the one shown in Fig. 3.3.4.1. We had to make a consideration with the angle of cut, as the most optimum was "th = 40", but because with this value the overall size of the antenna was almost the same, a compromise between S11 and dimension is made, so we choose the 30 degree angle.

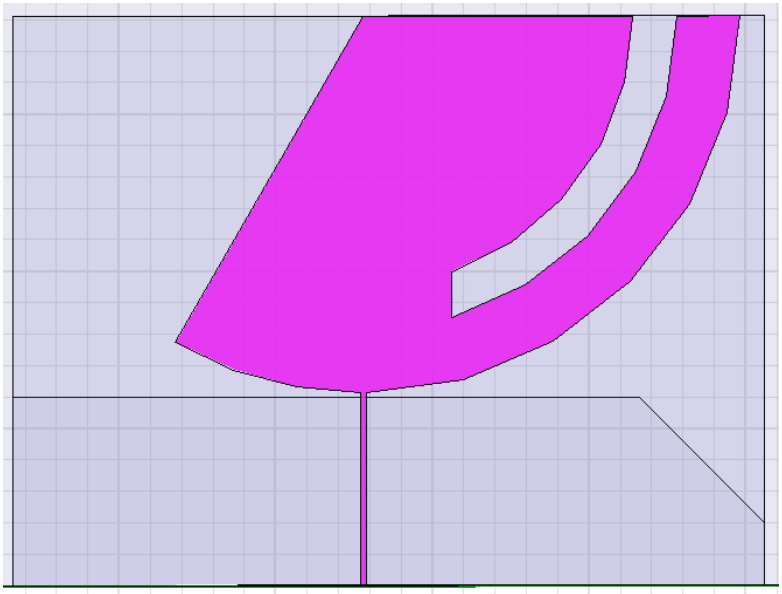


Fig. 3.3.4.1: Size-Improved antenna.

Table 3.3.2.1: Final values for the 2D base-antenna, the name of the values are taken from Fig. 3.3.3.1 and Fig. 3.3.3.6

Name	Value
L	90.6 mm
W	120 mm
l	20 mm
gr	30 mm
g	0.6 mm
r	60 mm
dd	14 mm
dr	10mm
dw	7mm
th	30 degrees
o	8mm

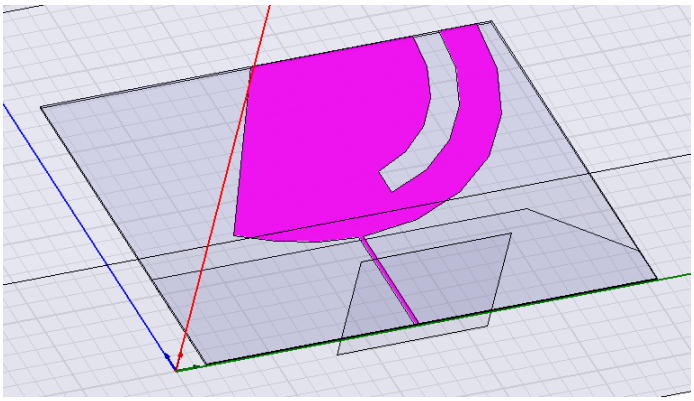
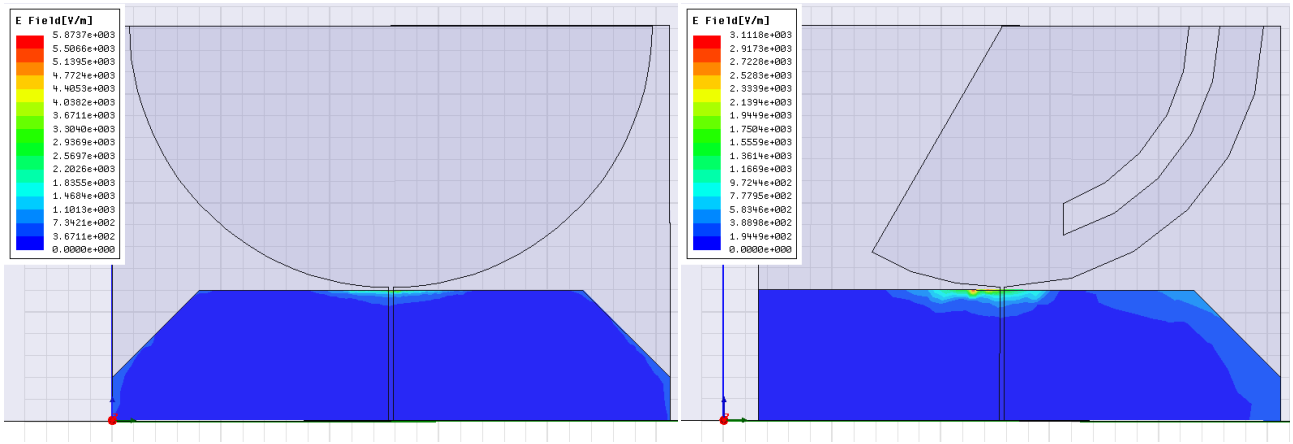
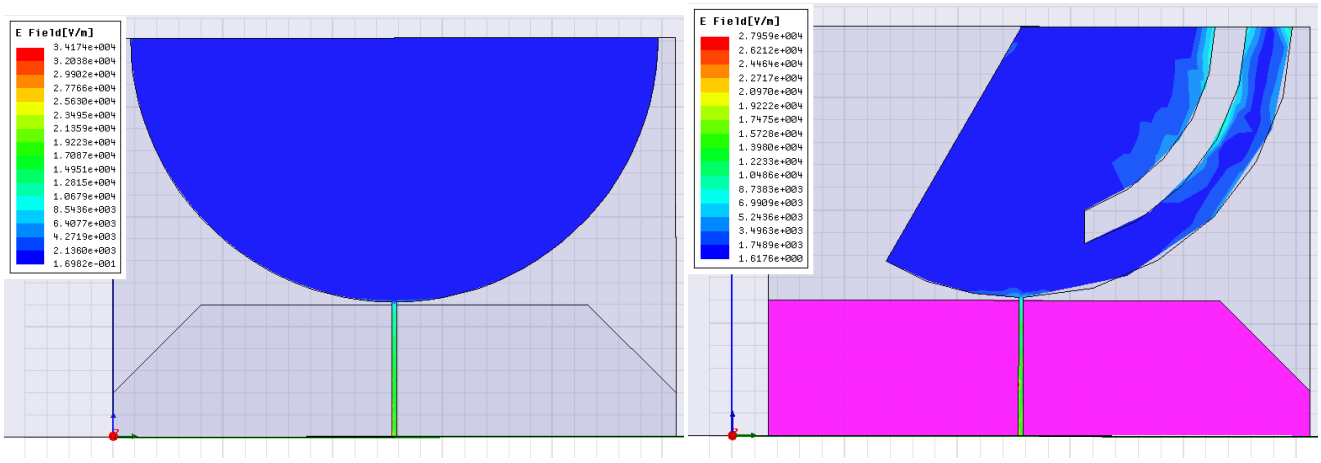


Fig. 3.3.4.2: 3D view of the size-improved antenna (HFSS11)

As explained above, now that we have decided on our new size-improved antenna values, the current distribution between the antennas are not the same. In the following figures we will appreciate that, as for the radiating patch and for the ground plane as well, the current distributions have changed. The differences for each frequency could also be seen clearly.

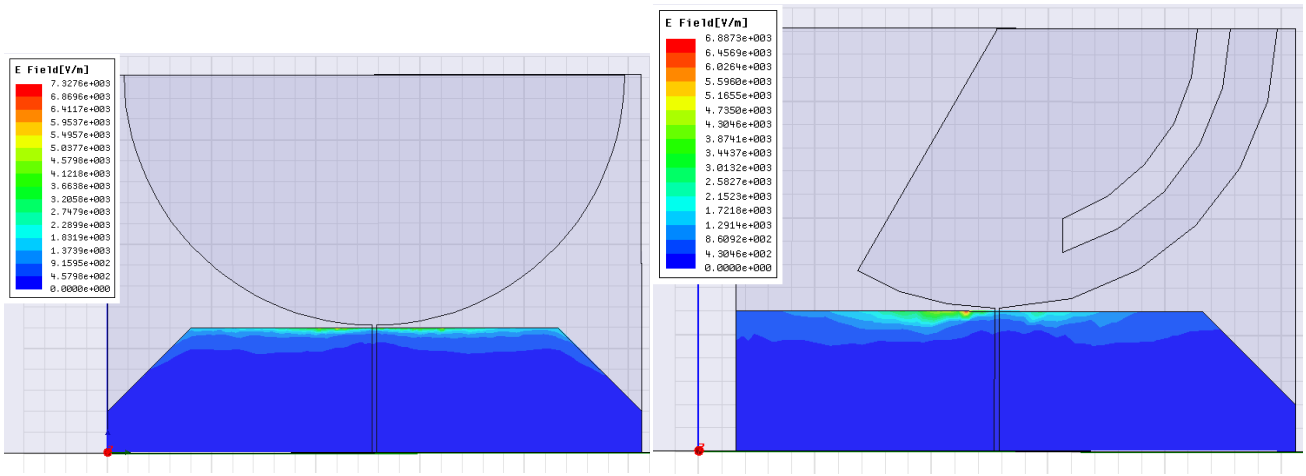


(a)



(b)

Fig. 3.3.4.3: Current distributions for (a) ground plane and (b) radiating patch at 900 MHz



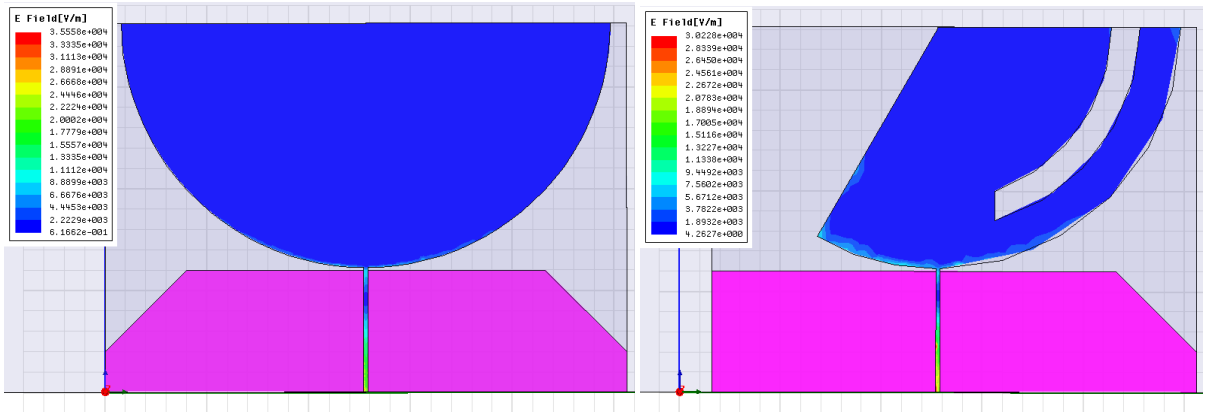


Fig. 3.3.4.4: Current distributions for (a) ground plane and (b) radiating patch at 1.85 GHz

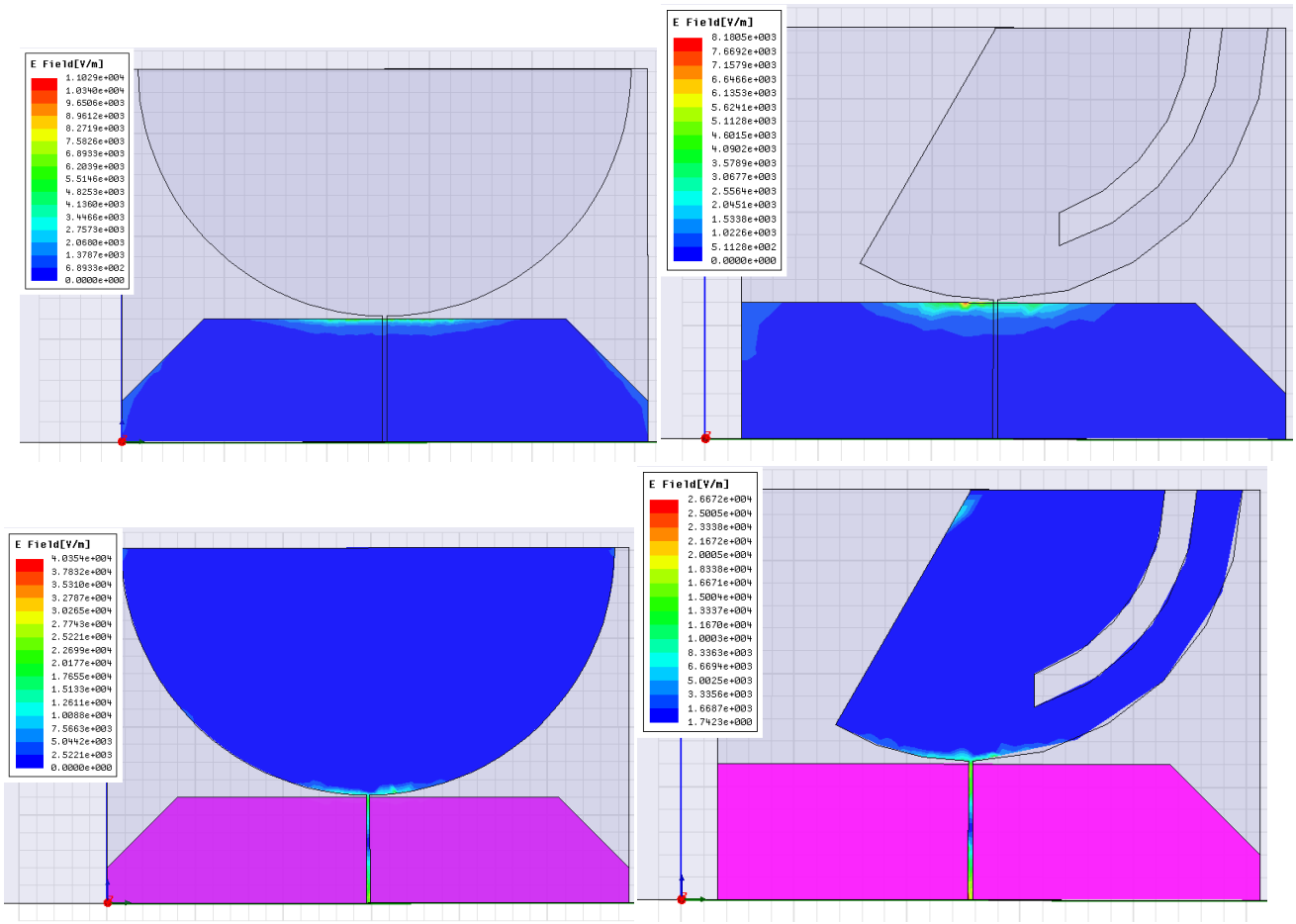
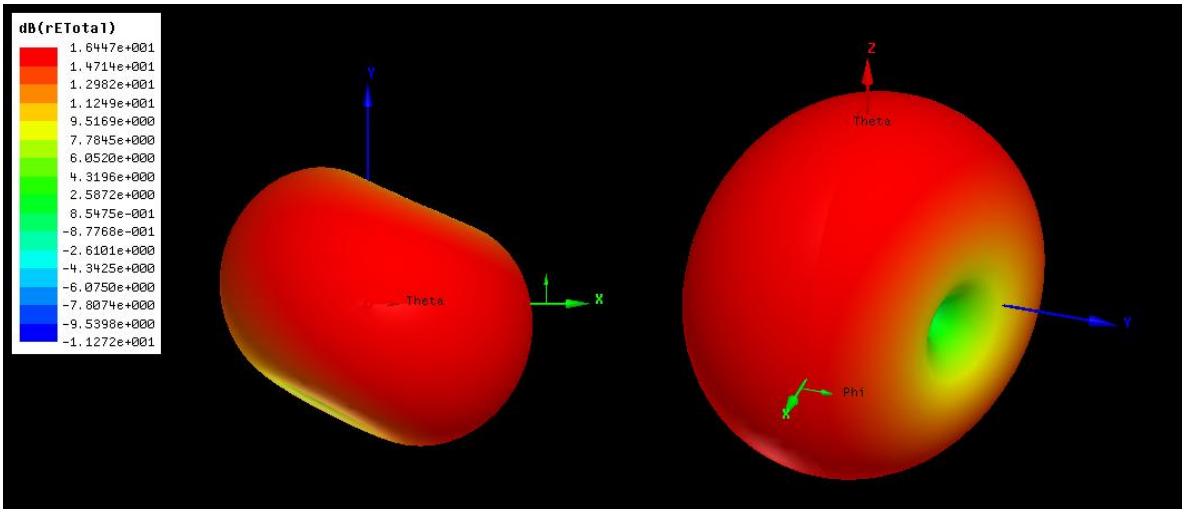


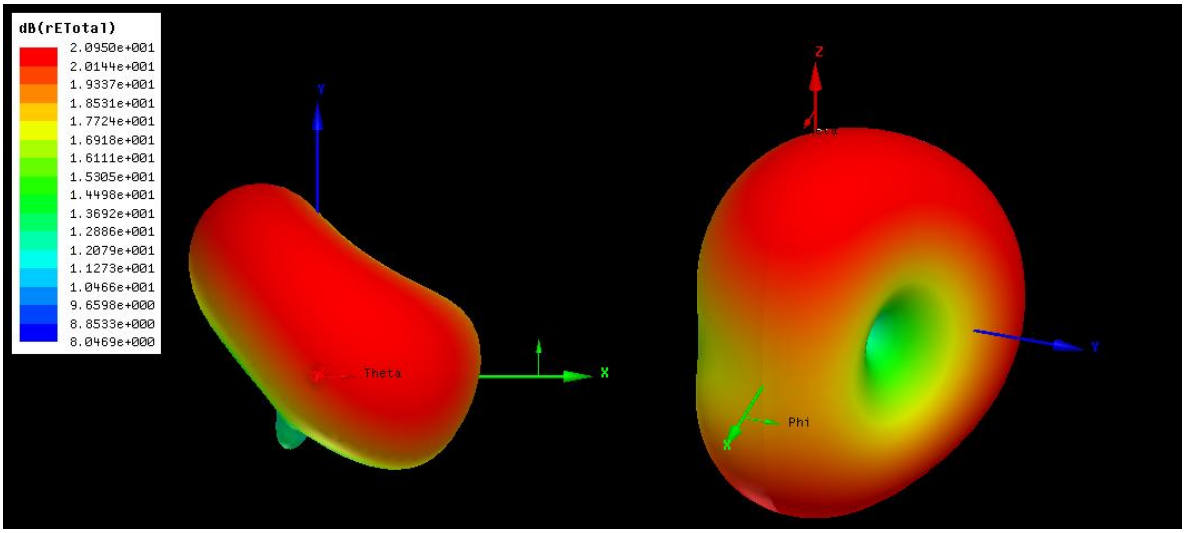
Fig. 3.3.4.5: Current distributions for (a) ground plane and (b) radiating patch at 2.45 GHz

We can see that for low frequencies, most of the current is concentrated around the right side of the antenna, where the slot is located. And we also can observe that the current on the left side of the antenna is very weak. The graphs also suggest that this behavior (current distribution) affects lower frequencies more than higher frequencies. For higher frequencies, we see that the current is distributed in the feed line and the base of the antenna.

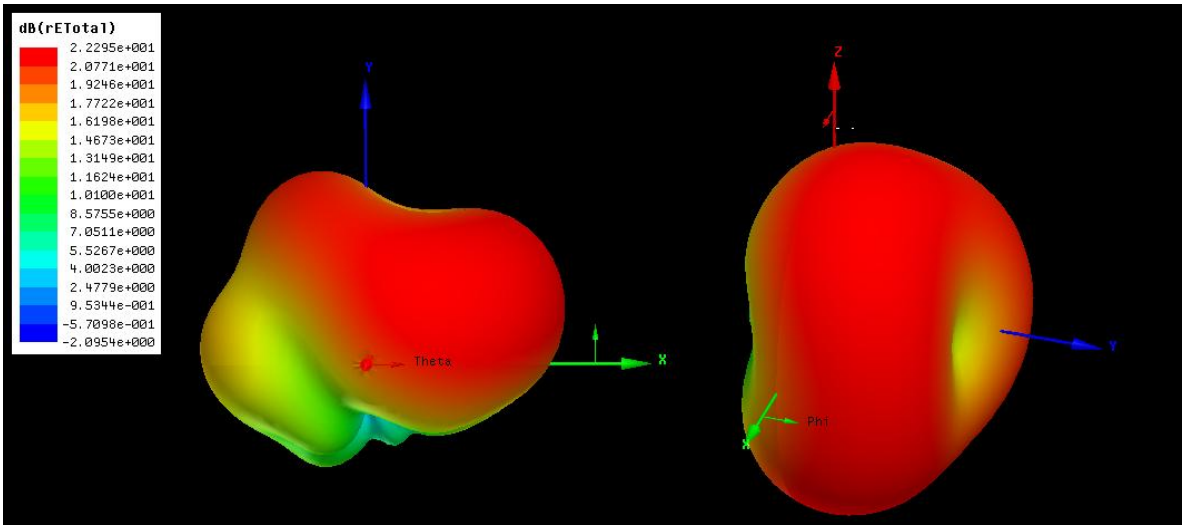
Some of the drawbacks of applying these techniques miniaturization are that the result, as in our case, turns out to be asymmetrical. The feed lines, as well as the notch in the ground plane are no longer symmetrical. Direct consequence of this asymmetrical physical change is the radiation pattern.



(a)



(b)



(c)

Fig. 3.3.4.6: Size-improved antenna radiation pattern at(a) 900 MHz, (b) 1.85 Ghz and (c) 2.45 GHz

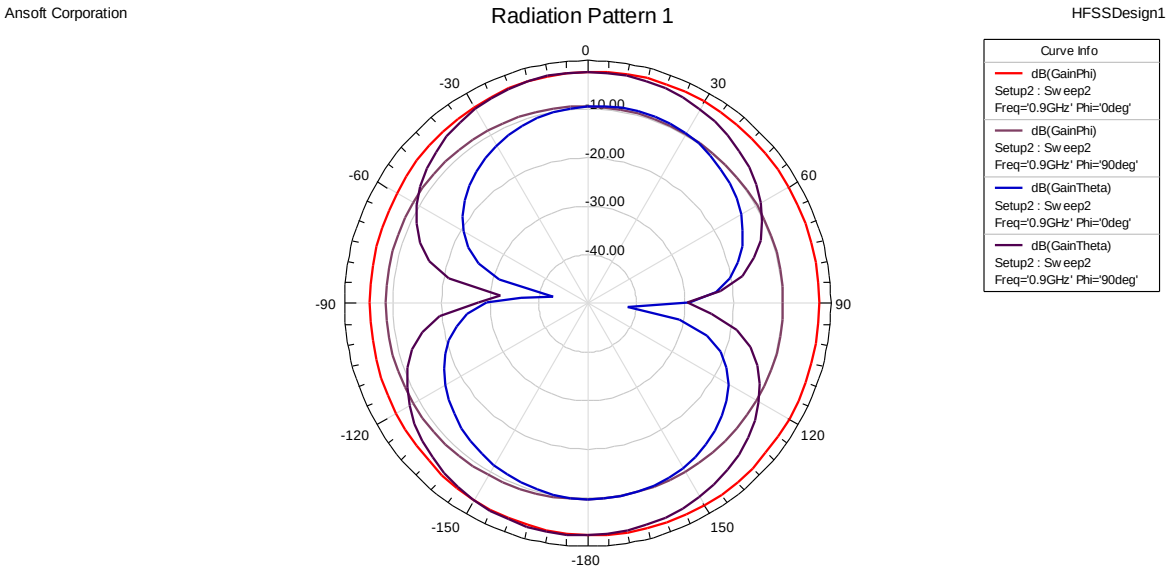


Fig. 3.3.4.7: Size-improved antenna gain pattern at 900 MHz

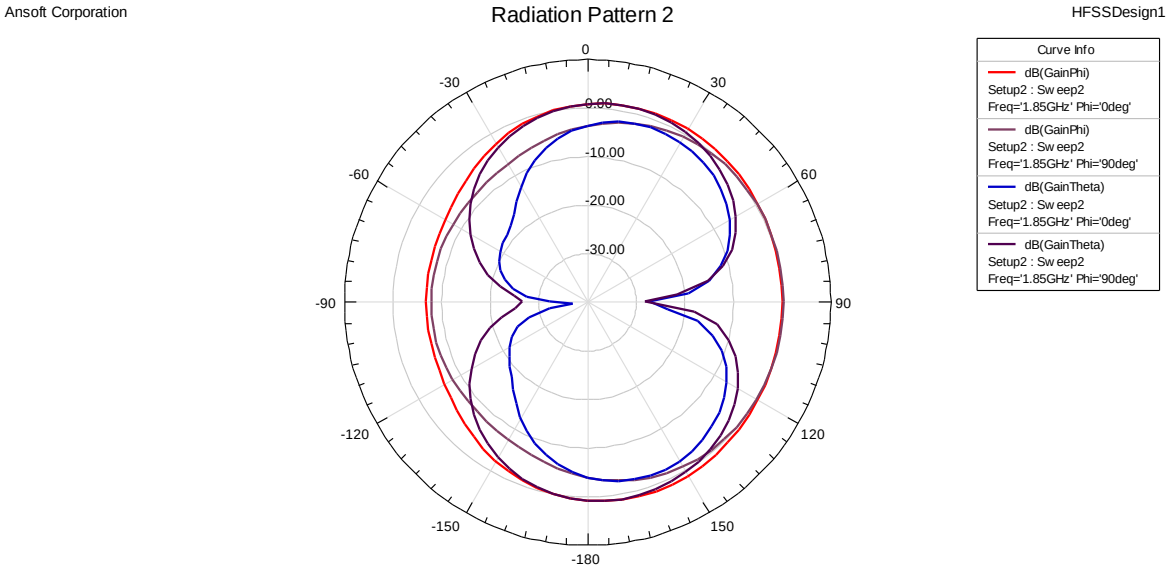


Fig. 3.3.4.8: Size-improved antenna gain pattern at 1.85 GHz

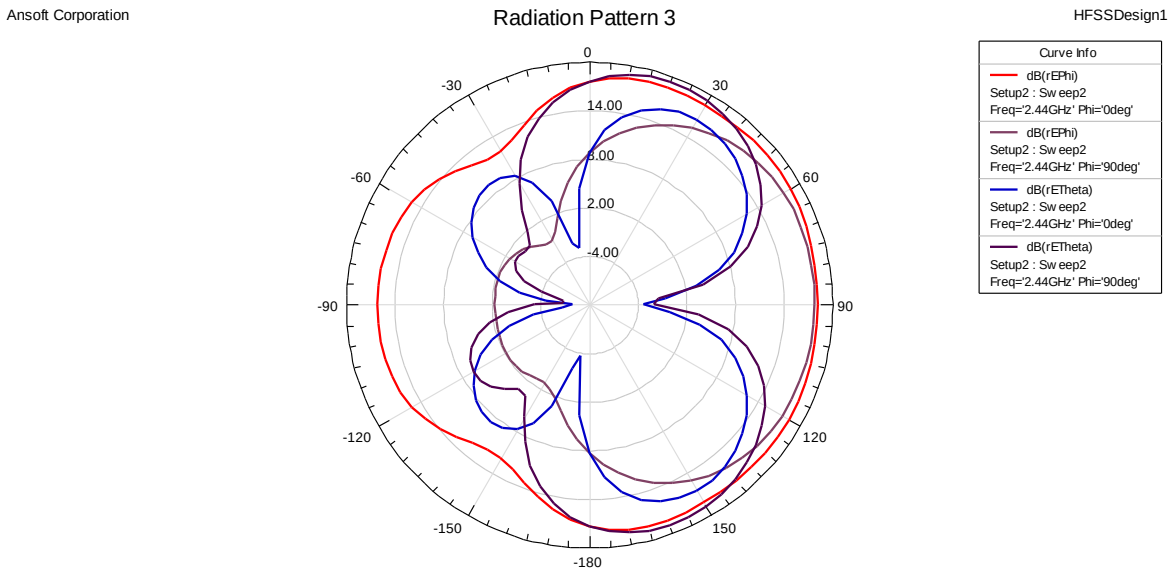


Fig. 3.3.4.9: Size-improved antenna gain pattern at 2.45 GHz

3.4 Antenna Fabrication and measurements

3.4.1 Fabrication

For the manufacture of the antennas, a LPKF Protomat C100 HF Milling machine (Fig. 3.4.1.1) was used. This machine has a set of drills to fit the milling desired size (Fig. 3.4.1.2). First we exported the antenna design file from the HFSS11 software to the software used by the machine. This machine cuts the metal layer of the used substrate (A25N) so that we "draw" the antenna shape (Fig 3.4.4.2). After having the shape of the antenna, we have to turn the substrate to repeat the same procedure to "draw" the ground shape. After it, all the substrate is separated from the entire substrate plate by making a deeper cut. Then, using a hand cutter, we remove the exceeding copper layers (Fig. 3.4.1.4) that are not part of our antenna design. Finally, we have to solder a 3.5mm SMA female connector on the fed line.

Fig 3.4.1.1: Milling machine

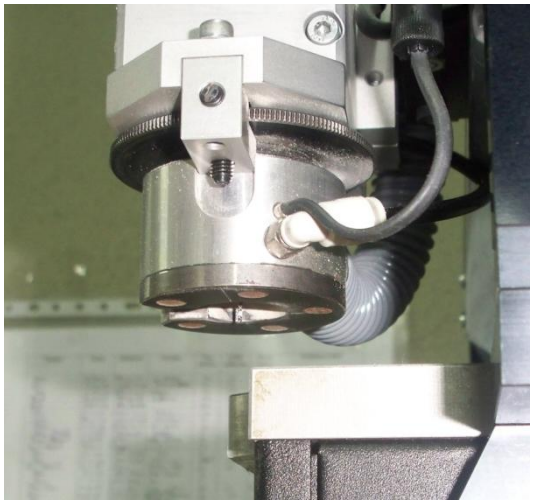


Fig 3.4.1.2: Cutting Drill

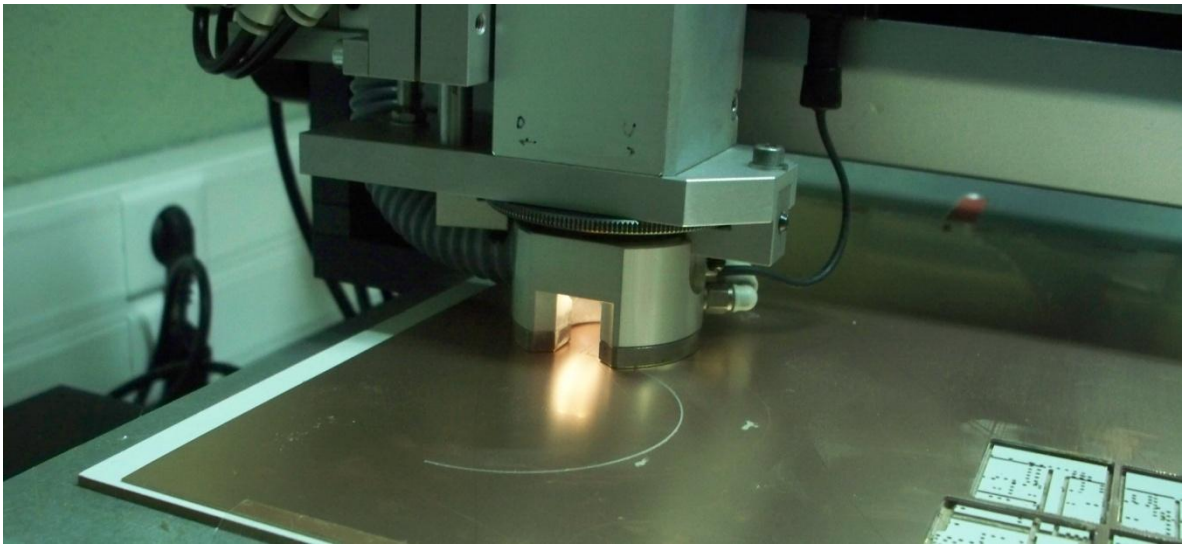


Fig 3.4.1.3: Milling machine shaping our antenna

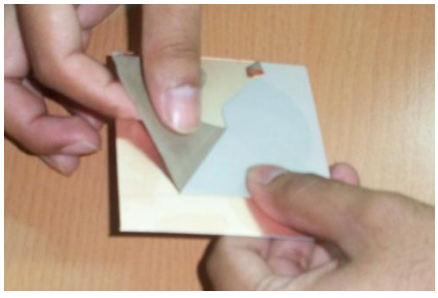


Fig 3.4.1.4: Exceeding Cupper

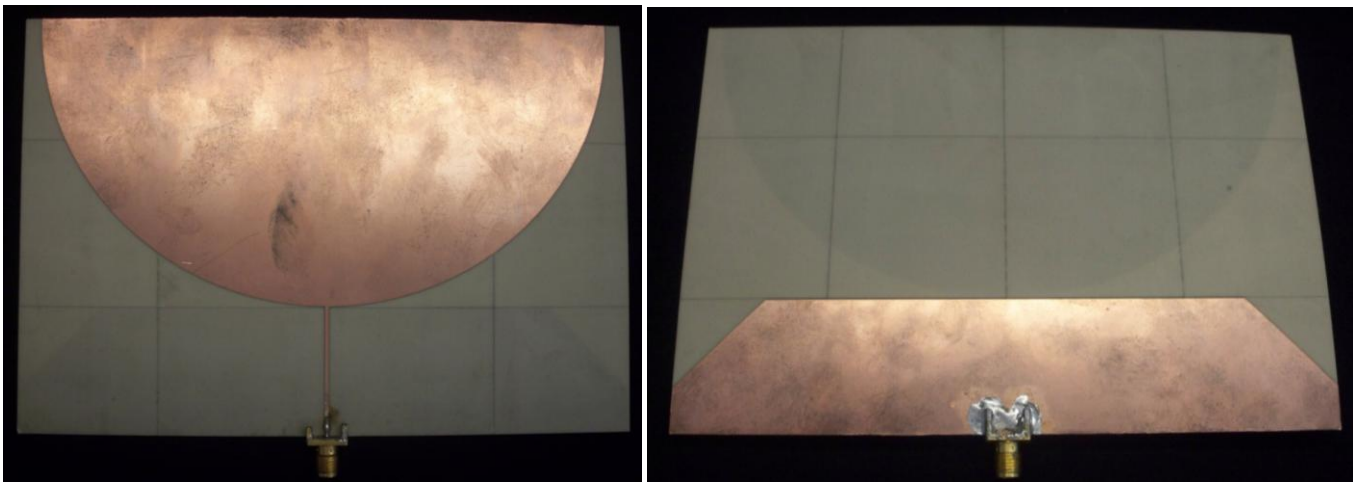


Fig. 3.4.1.5: base-antenna front and back view

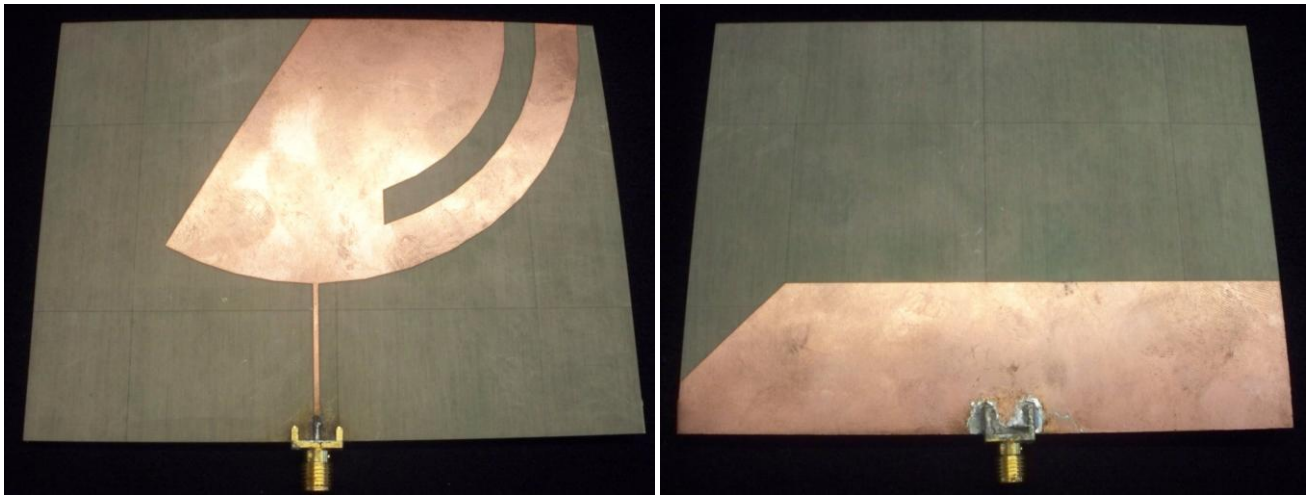


Fig. 3.4.1.5 size-improved antenna front and back view

3.4.2 Measurements

For measurements of the antennas was used a ZVA24 Rohde & Schwarz Vector Network Analyzer (VNA). Since we want to know the energy that our antenna "bounces" and these frequencies, then we have to measure the S11 parameter. A deeper explanation of the parameters S in section 4.3.1 will be done.

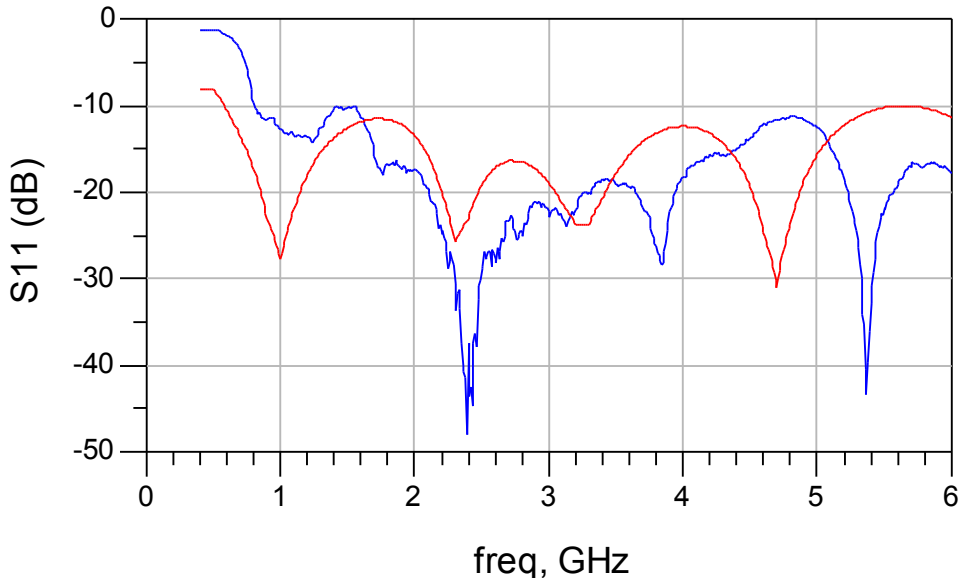


Fig. 3.4.2.1: S11 parameters measured for the base antenna in red compared with the simulated results in blue.

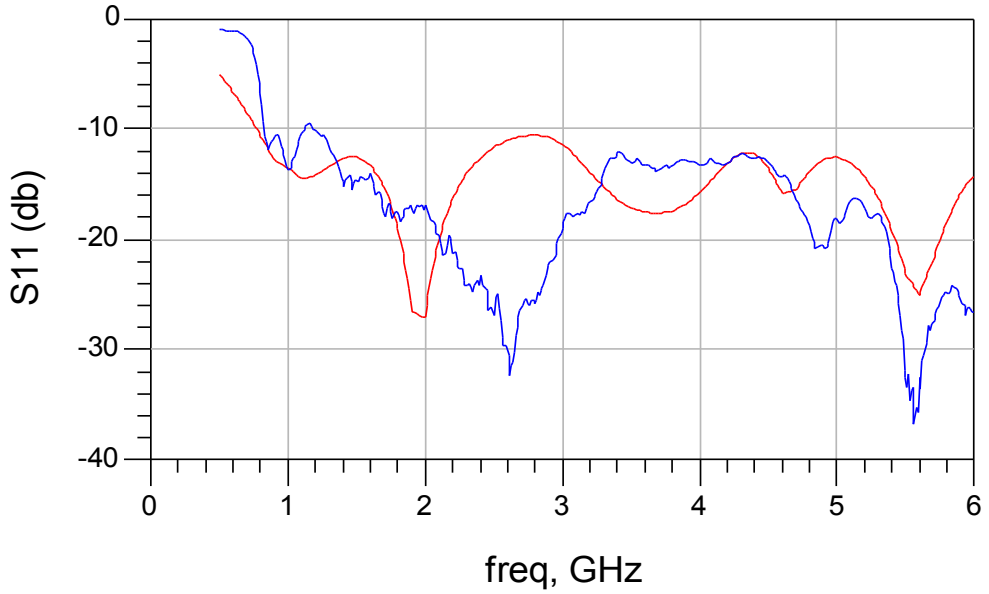


Fig. 3.4.2.2: S11 parameters measured for the size-improved antenna in red compared with the simulated results in blue.

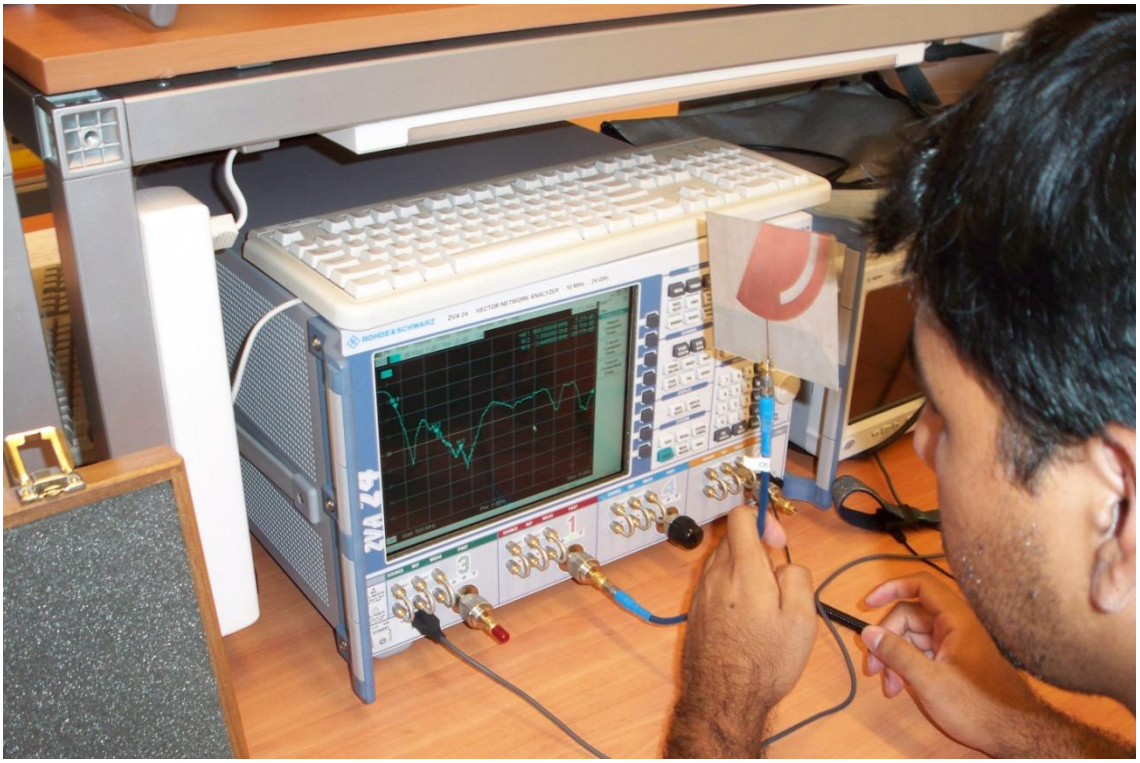


Fig. 3.4.2.3 Vector Network Analyzer, making some measurements.

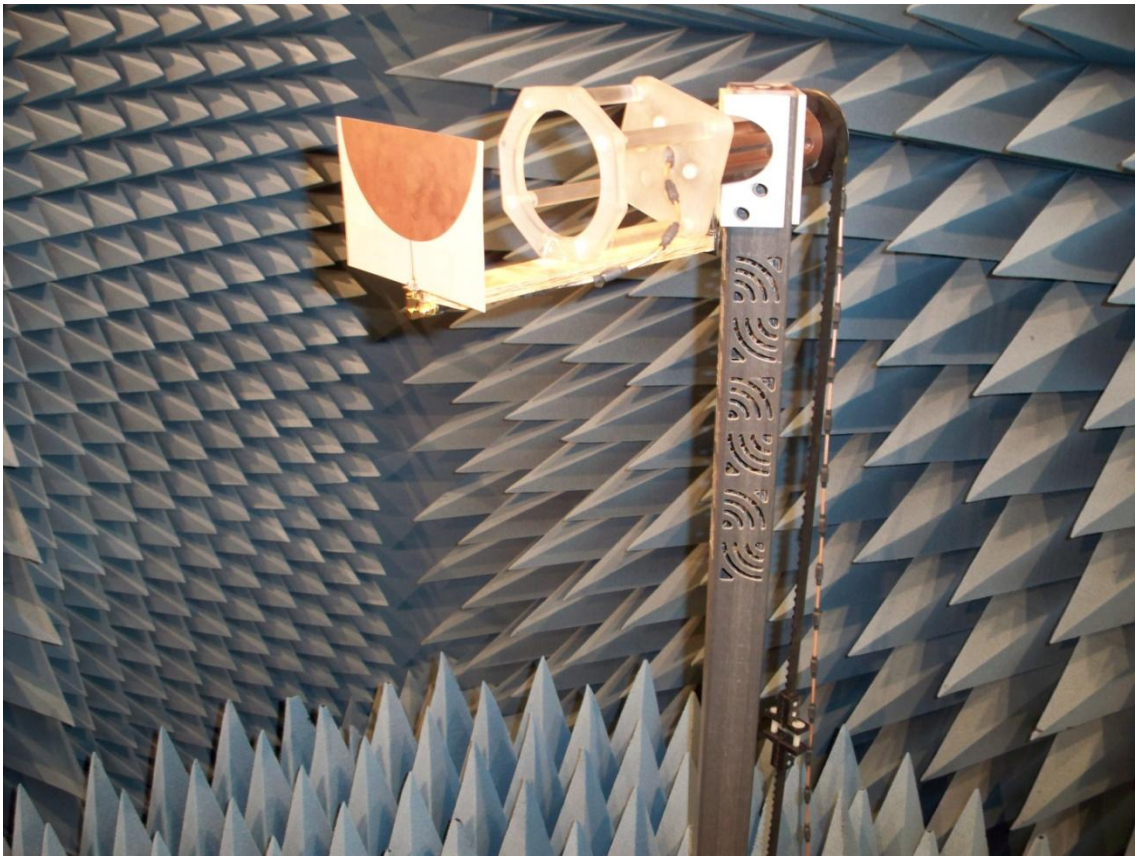


Fig. 3.4.2.4: Anechoic Chamber.

Fig. 3.4.2.1 and Fig. 3.4.2.2 compare the VNA measurements with th software simulations. We can conclude that there is a good correspondence between the measurements and simulations.

Measurements in the anechoic chamber are used for far field antenna characteristics. In Figs. 3.4.2.5 – 3.4.2.8, that show the radiation patterns, rETheta (φ = 0) and rEPhi (φ = 90), are cross-polarization measurements and rETheta (φ = 90) and rEPhi (φ = 0), represent co-polarization measurements. It can be seen that in Figure 3.4.2.5 and 3.4.2.6, which are measurements of the base antenna; there is a marked polarization in the cross-polar direction, and in Fig. 3.4.2.7 and Fig. 3.4.2.8 (Size-improved antenna), theoretically with similar behavior than the first antenna, the measurements show some response in the cross-polar direction.

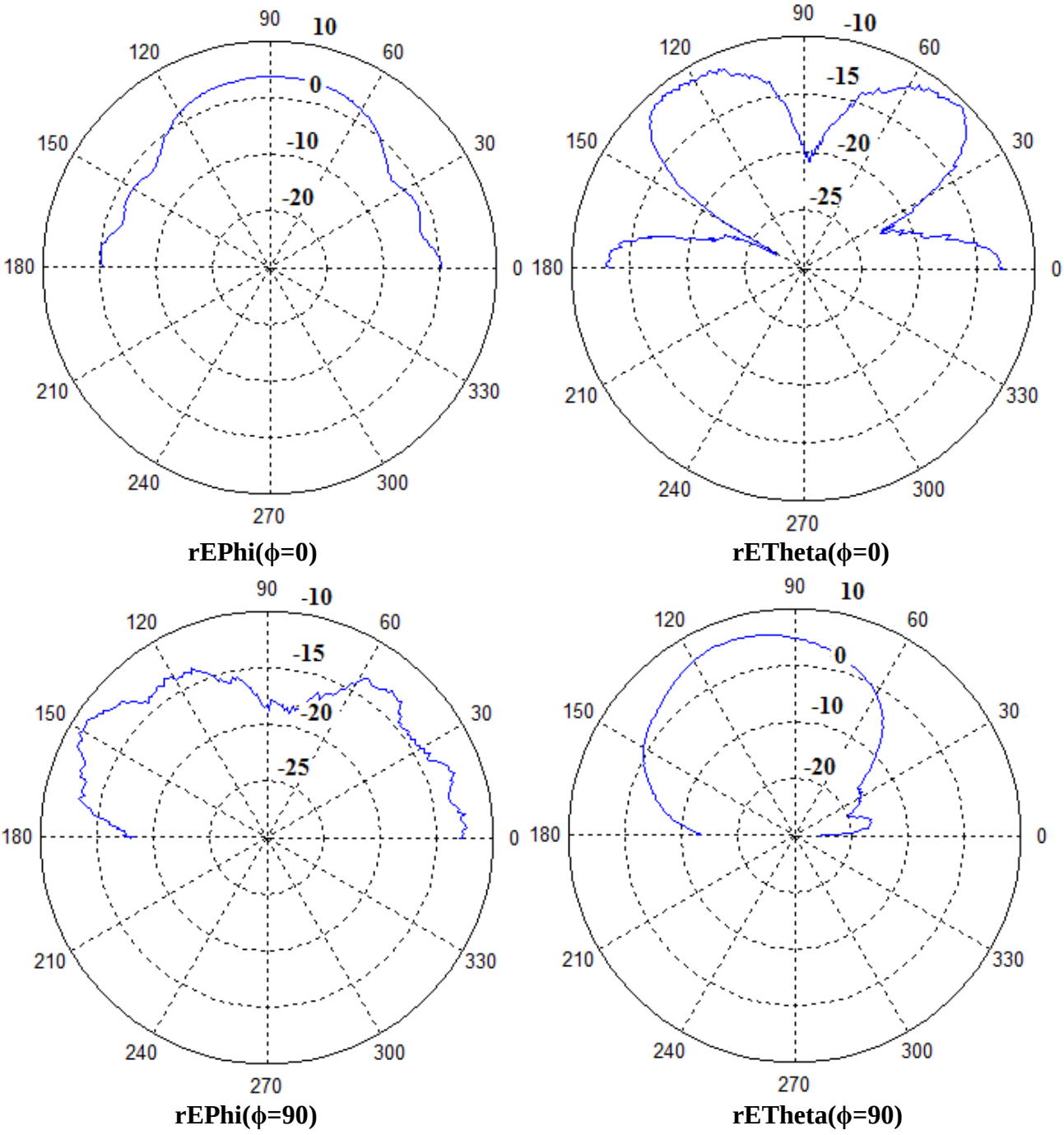


Fig. 3.4.2.5 measured radiations patterns at 1.85 GHz for the base-antenna.

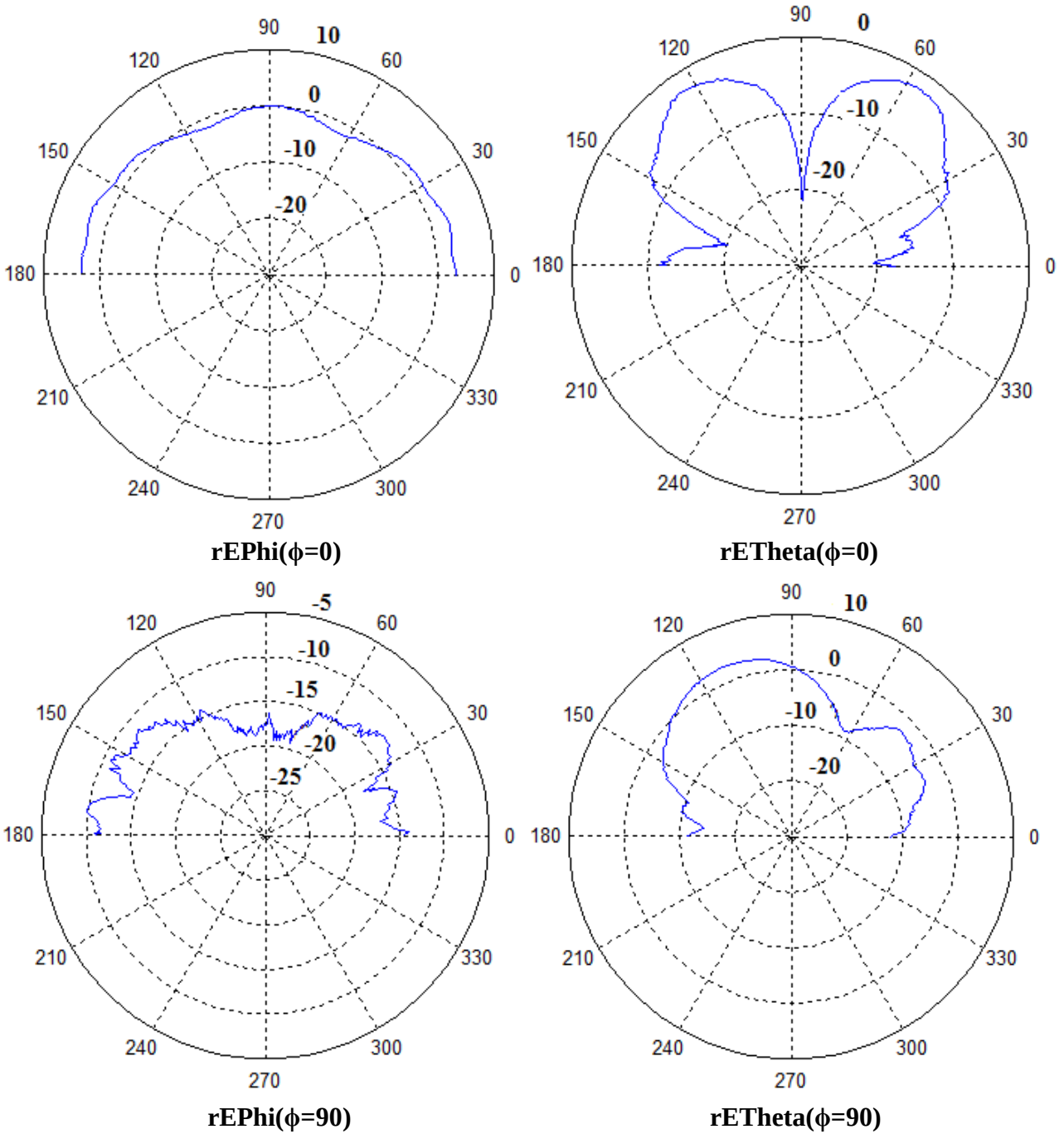


Fig. 3.4.2.6 measured radiations patterns at 2.45 GHz for the base-antenna.

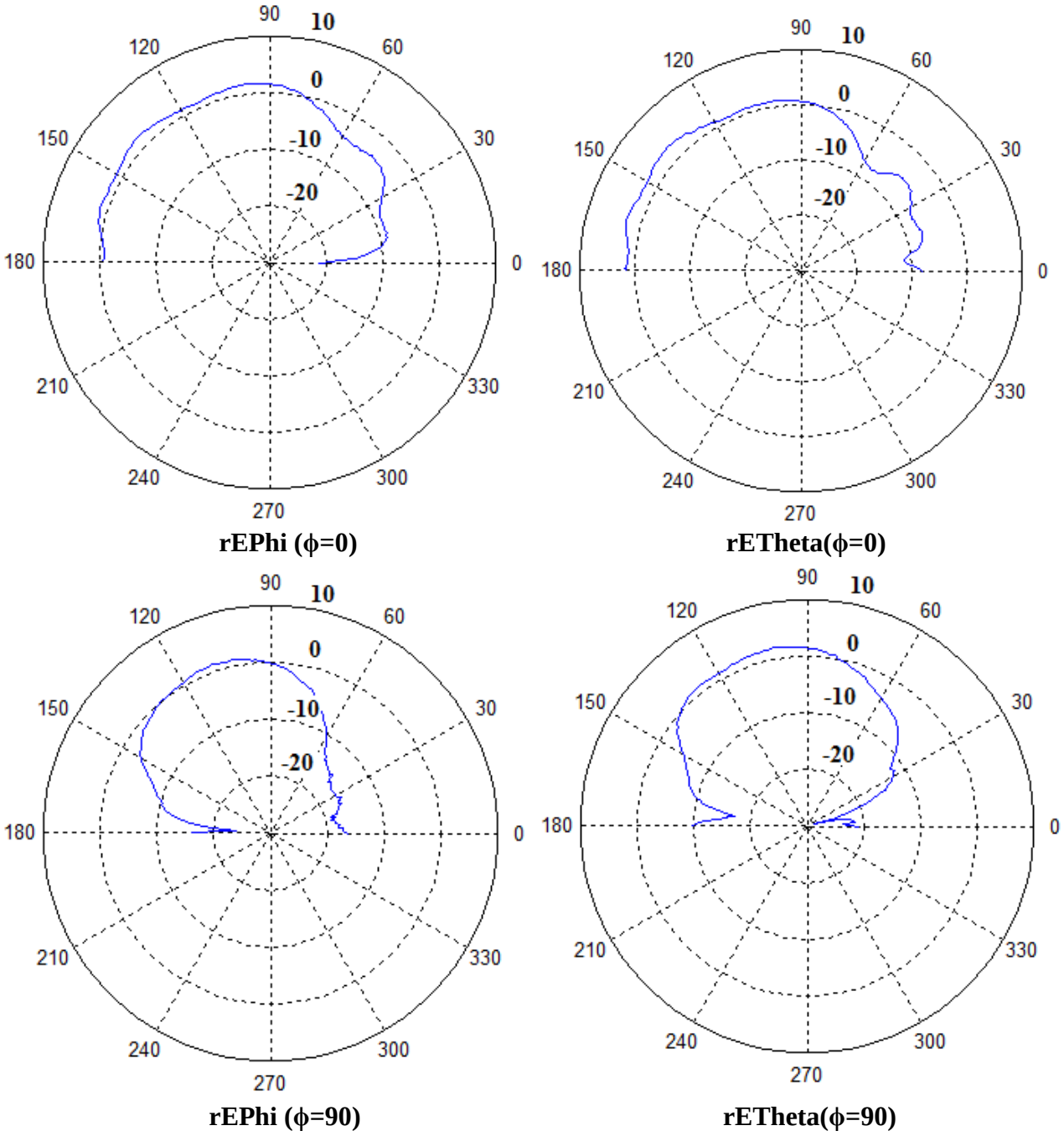


Fig. 3.4.2.2 measured radiations patterns at 1.85 GHz for the size-improved antenna.

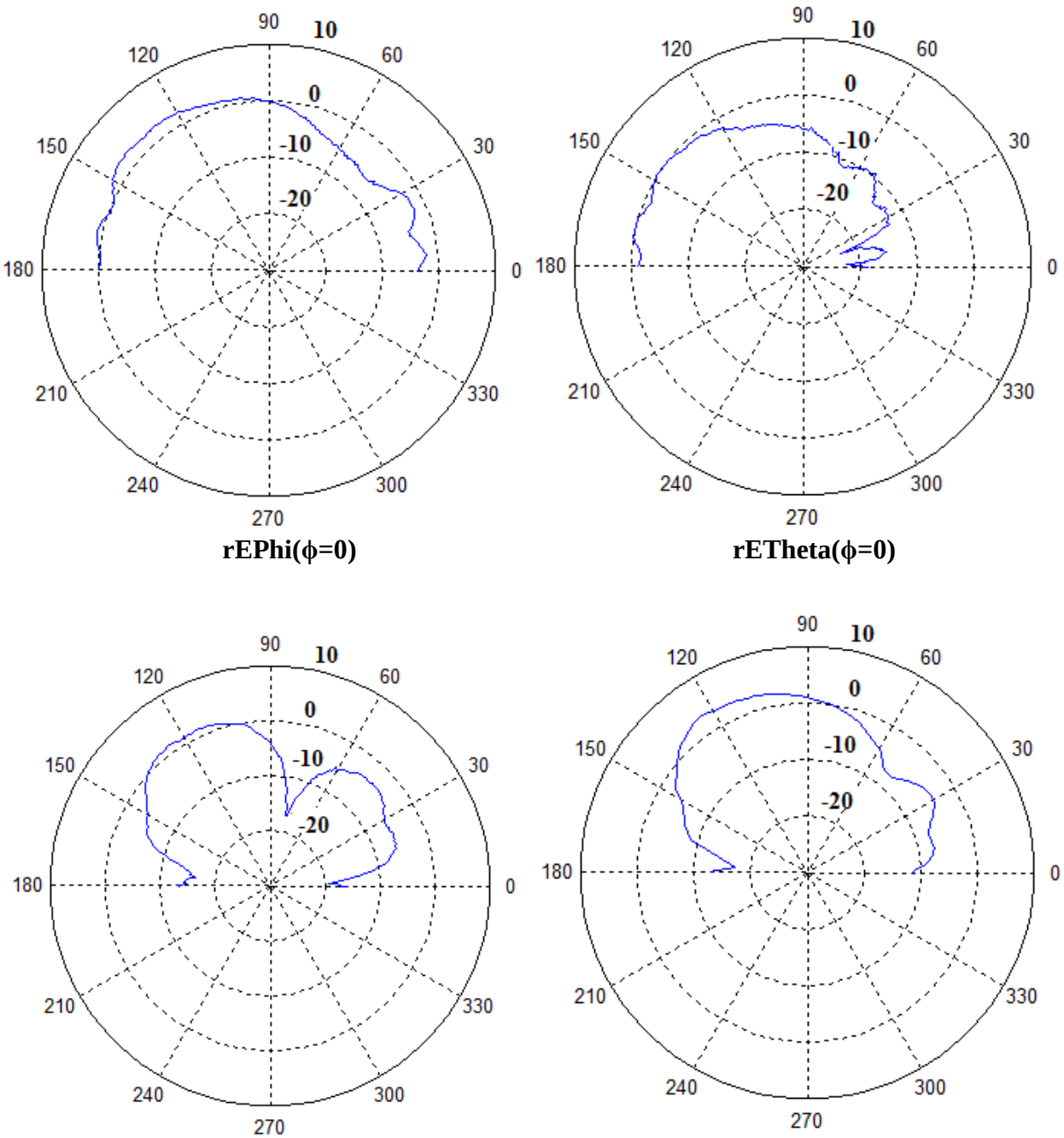


Fig. 3.4.2.8 measured radiations patterns at 2.44 GHz for the size-improved antenna.

Chapter 4

Rectifier Design

4.1 Rectifier

A rectifier is an electrical device that converts alternating current (AC), which periodically reverses direction, to direct current (DC), current that flows in only one direction, in a process known as rectification. Rectifiers have many uses including as components of power supplies and as detectors of radio signals.

There are two types of rectifier, the half wave and full wave. Each type can either be uncontrolled, half-controlled or fully controlled. An uncontrolled rectifier uses diodes, while a full-controlled rectifier uses thyristor or popularly known as Silicon Controlled Rectifier (SCR). A half controlled is a mix of diodes and thyristors. The thyristors need to be turned on using a special triggering circuit. In our application, given the ease of matching because of the fewer number of components, has been used a half-wave rectifier.

4.1.1 Commutating Diode

Mostly, for half or uncontrolled circuit they include also a diode across the load as shown in Fig. 4.1.3.1 This diode is variously described as a freewheeling diode (FWD), flywheel or bypass diode, but is best described as a commutating diode as its function is to commutate or transfer load current away from the rectifier whenever the load current away from the rectifier whenever the load voltage goes into a reverse state.

4.1.2 Schottky diode

The Schottky diode is a semiconductor diode with a low forward voltage drop and a very fast switching action. A Schottky diode is a special kind of diode with a very low forward-voltage drop. When current flows through a diode there is a small voltage drop across the diode terminals. A normal silicon diode has between 0.6–1.7 volt drops, while a Schottky diode voltage drop is between approximately 0.15–0.45 volts. This lower voltage drop can provide higher switching speed and better system efficiency. For our rectifier, we have used a SKYWORKS® Schottky diode (Fig. 4.1.2).

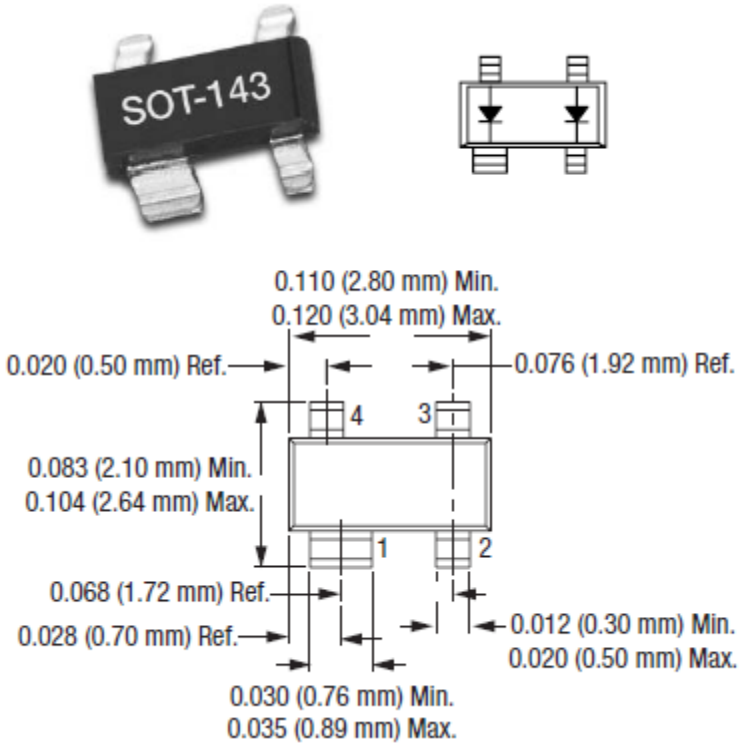


Fig. 4.1.2 SKYWORKS Schottky diode picture, pins and dimensions

4.1.3 Half-Wave Rectifier

In practice, the half-wave rectifier is used most often in low-power applications because the average current in the supply will not be zero. This may cause problems in transformer performance.

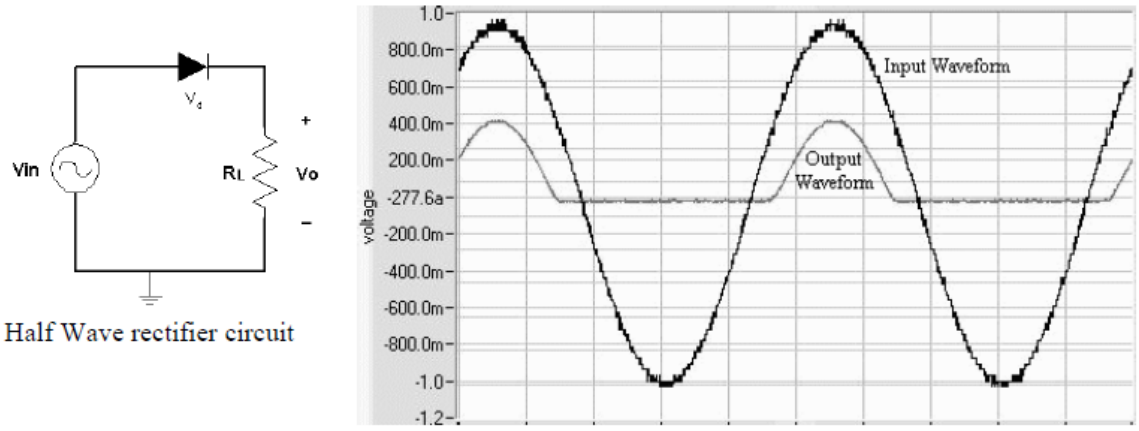


Fig. 4.1.3.1 Half wave rectifier input and output wave example

When a capacitor is connected in parallel with RL, the half wave output becomes a "filtered" DC. As the voltage of the input sine wave increases, the diode will begin to conduct and charge the capacitor. When the input sine wave voltage falls below the capacitor voltage, the diode turns off and the capacitor discharges into the load.

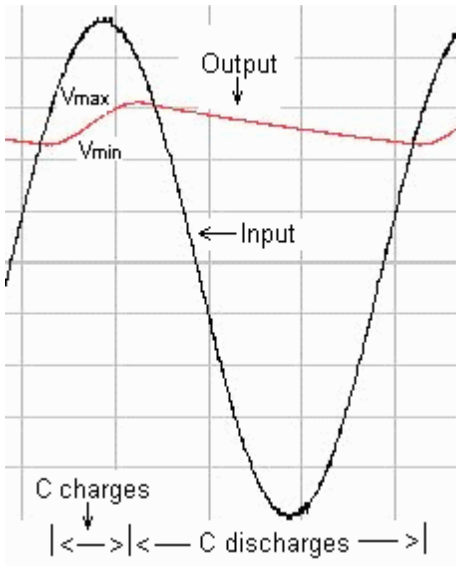


Fig. 4.1.3.2: Operation of the rectifier when placing a capacitor in parallel with RL

The size of the resistor will determine Vmin. As the load resistance is increased: IL decreases, the capacitor discharges less, Vmin increases, Vrms (effective voltage) increases, and Vpp (ripple voltage) decreases. Vmax does not change much because the diode forward voltage drop is relatively constant and Vmax = Vin – Vd.

4.2 RF to DC efficiency

At some applications explained in chapter two and others like the reception of a microwave beam sent to the earth by a geostationary solar power satellite, distributed DC powering of actuator, or DC supplying of RFID passive tags, the RF to DC conversion efficiency is the parameter to improve over the range of incident power or output voltage of interest. Different solutions have been reported, mainly based on combining dipole or patch printed elements with diode detectors. The RF-DC conversion efficiency of the rectenna with a diode depends on the microwave power input intensity and the connected load. We need to get the optimum microwave power input intensity and the optimum load to achieve maximum efficiency. When the power or load is not well matched, the transferred power is not the optimal and the efficiency becomes quite low. The efficiency is also determined by the characteristic of the diode; the diode has its own junction voltage and breakdown voltage. If the input voltage to the diode is lower than the junction voltage or is higher than the breakdown voltage, the diode does not show a rectifying characteristic. As a result, the RF-DC conversion efficiency drops with a lower or higher input than the optimum.

The average RF power over a range of frequencies at any instant in time is given by:

$$\left(\frac{1}{f_h - f_l}\right)\int_{f_l}^{f_h}\int_{0}^{4\pi} S(\theta,\phi,f,t)A_{eff}(d\Omega df)$$

The DC power for a single frequency (fi) input RF power is given by:

$$P_{DC}(f_i) = P_{RF}(f_i,t)\eta[P_{RF}F(f_i,t),\rho,Z_{DC}]$$

where η is the conversion efficiency:

$$\eta[P_{RF}F(f_i,t),\rho,Z_{DC}] = \frac{P_{DC}(f_i)}{P_{RF}(f_i,t)}$$

4.3 Rectifier Simulation Tools

For rectifier simulations and also each of its components, we have used the Agilent ADS2008 software. In this section we describe some of the modules and tools used, and a brief explanation of its operation is also made.

4.3.1 S-parameter analysis.

The objective of microwave circuit analysis is to move from the requirement to solve all the fields and waves of a structure, to an equivalent circuit that is amenable to all the tools of the circuit analysis. However, the tools that are appropriate for lumped circuits must be extended to apply to distributed networks.

A matrix that is of great use in microwave network problems is the "scattering" matrix, so-called by analogy to the scattering or reflection of waves by a free-space reflector. S-parameters have become the preferred description of microwave n-ports for the following reasons:

- Voltage and current are difficult to define and measure in distributed circuits
- Incident and reflected waves are the natural description for microwave structures
- Conversion from S-parameters to other parameter sets is a matter of routine algebra
- S-parameters benefit from the matrix operations.

S-parameter simulation is a type of small-signal AC simulation. It is most commonly used to characterize a passive RF component and establish the small-signal characteristics of a device.

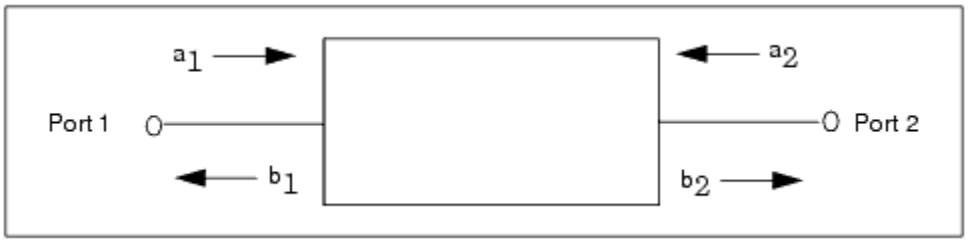


Fig. 4.3.1.1: Representation of a signal wave in a two-port electrical-element.

Fig. 4.3.1.1 shows a representation of a signal wave in a two-port electrical-element where:

- *a1* is the wave into port 1
- *b1* is the wave out of port 1
- *a2* is the wave into port 2
- *b2* is the wave out of port 2

The S-parameters for this conventional element are:

- $b1 = a1.S11 + a2.S12$
- $b2 = a1.S21 + a2.S22$

where:

- S11 is the port-1 reflection coefficient: S11= *b1/a1* when *a2*= 0
- S22 is the port-2 reflection coefficient: S22=*b2/a2* when *a1*= 0
- S21 is the forward transmission coefficient: S21= *b2/a1* when *a2*= 0
- S12 is the reverse transmission coefficient: S12= *b1/a2* when *a1*= 0

These equations can be solved for *b1* and *a1* in terms of *a2* and *b2* to yield the transmission T-parameters as follows:

- *b1*= *a2*.T11+ *b2*.T12
- *a1*= *a2*.T21+ *b2*.T22

Fig. 4.3.1.2 shows the T-Parameters matrix.

$$\begin{bmatrix} t_{11} & t_{12} \\ t_{21} & t_{22} \end{bmatrix} = \begin{bmatrix} s_{12} - s_{11} & s_{22}/s_{21} \\ -s_{22}/s_{21} & 1/s_{21} \end{bmatrix} \quad s_{11}/s_{21}$$

Fig. 4.3.1.2: relationship between the T-parameters and the S-parameters

S-parameters are defined with respect to reference impedance that is typically 50 ohms. For 50-ohm S-parameters-with the 2-port element terminated with 50 ohms at each port, the s21 parameter represents the voltage gain of the element from port 1 to port 2.

4.3.2 Harmonic Balance

Harmonic balance is a method to analyze the distortion in non-linear circuits or systems. This method is usually applied when we need to simulate RF or microwave problems since these are most naturally handled in the frequency domain.

Source voltage and current of a circuit can generate discrete frequencies by generating a discrete frequency spectrum at different points in the system. Linear components of a circuit are modeled in the frequency domain while non-linear components are modeled first in the time domain and then Fourier-transformed in the frequency domain before each solving step.

This method first divides the circuit into two groups, the linear elements, where are also included input impedances of the sources, and the group of non-linear elements. Fig. 4.3.2 shows a schematic of these two groups.

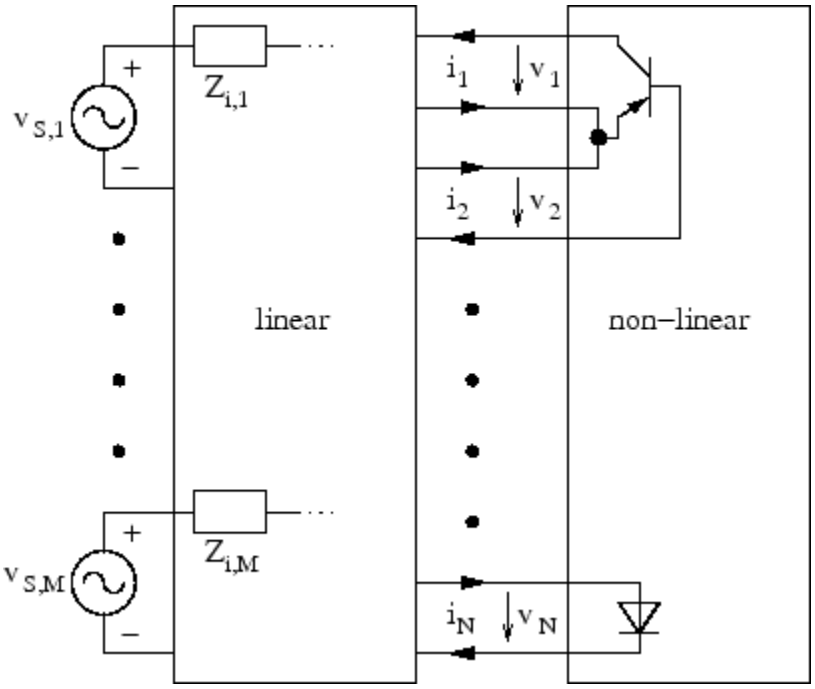


Fig. 4.3.2 Schematic of Harmonic Balance component division [20]

It is considered that a solution has occurred when the current between the groups is the same for the linear elements and for the non-linear. That is why the method is called harmonic balance, because the current between the two groups must be balanced for each harmonic frequency. It is also possible to know all the voltages at each of the interconnections between the two groups, once known these values of balanced voltages and currents, it performs a classic AC analysis and we can know the voltage and current value at any point of the circuit.

In our application we have used this method because the circuit has only one non-linear element that is the diode, moreover, the circuits that we have simulated does not require much computer processing, which makes the simulation faster.

4.3.3 ADS2008 Matching Utility

Since our application consists basically of an antenna and a rectifier circuit, is required an impedance matching to ensure good power transfer and performance. The ADS2008 impedance matching utility is a software tool that allows us to place a component in our scheme and, according to certain parameters manually defined, generate a matching circuit which can then be optimized according to our goals.

Fig. 4.3.3.1 shows an example of how the impedance matching component is placed in the circuit scheme. For our application we chose a band-pass matching circuit, that was

subsequently optimized for best performance in our application bands, because matching a network over a broad band is complex and it will require a larger amount of components.

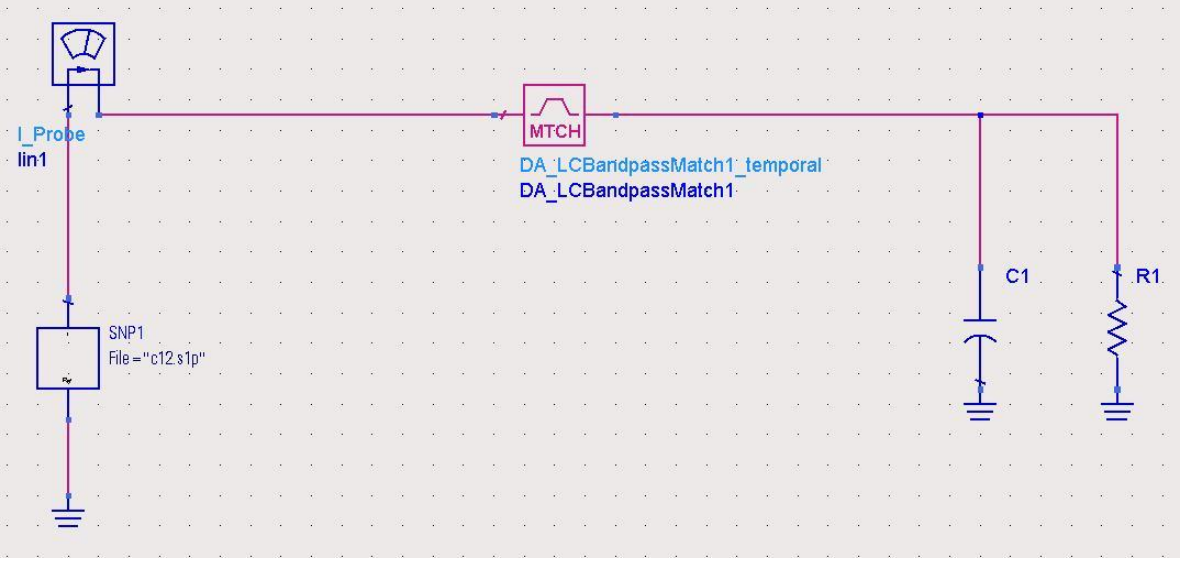


Fig. 4.3.3.1 Example circuit and Band pass matching component.

Although this component is placed in a circuit context, to generate the impedance matching circuit is not necessary that the component terminations are connected to any node, this is just done for further simulations, where the matching component replaces a whole (matching) circuit. When we deploy the menu of impedance matching (Fig. 4.3.3.1), we can enter the necessary parameters that will be implemented internally as well as the creation of the adaptation circuit.

Once we have defined our matching circuit, whose values were optimized according to our application bands, the next step is to optimize the whole set (adaptation circuit and rectifying circuit) to achieve our goals.

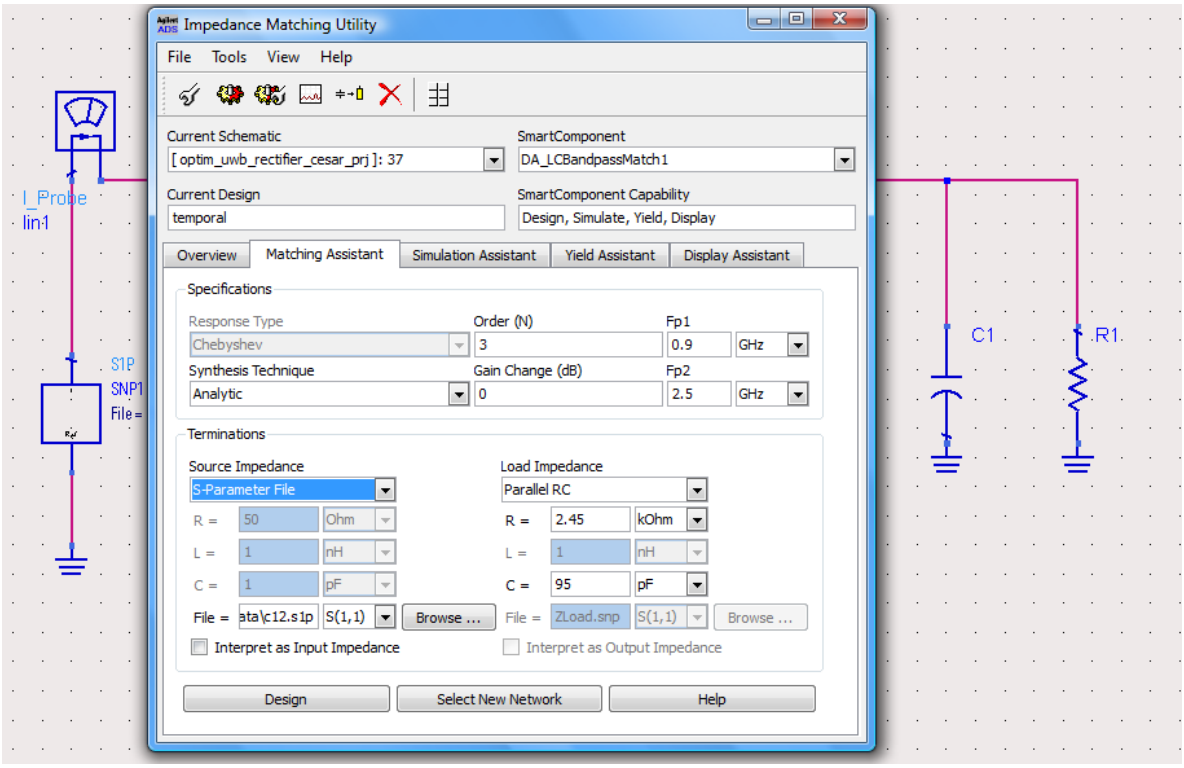


Fig. 4.3.3.2: Impedance matching assistant window. Main parameters to the filled: Lower Frequency (Fp1), Upper Frequency (Fp2). Source / Load Impedance

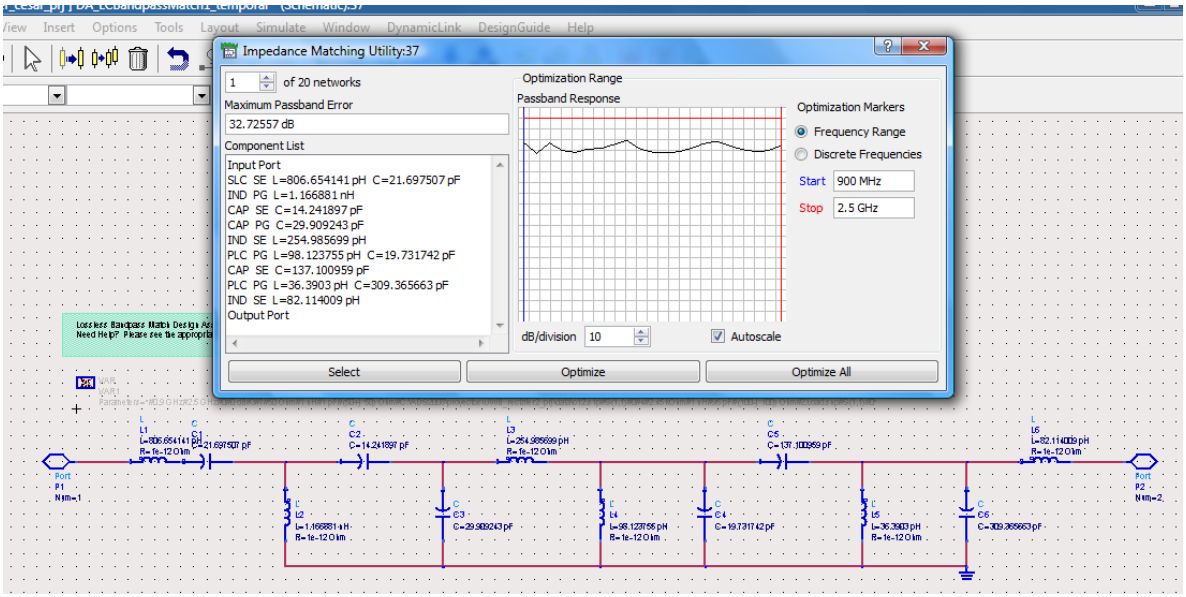


Fig. 4.3.3.3 automatically generated matching circuit. Note that there is more than one generated network.

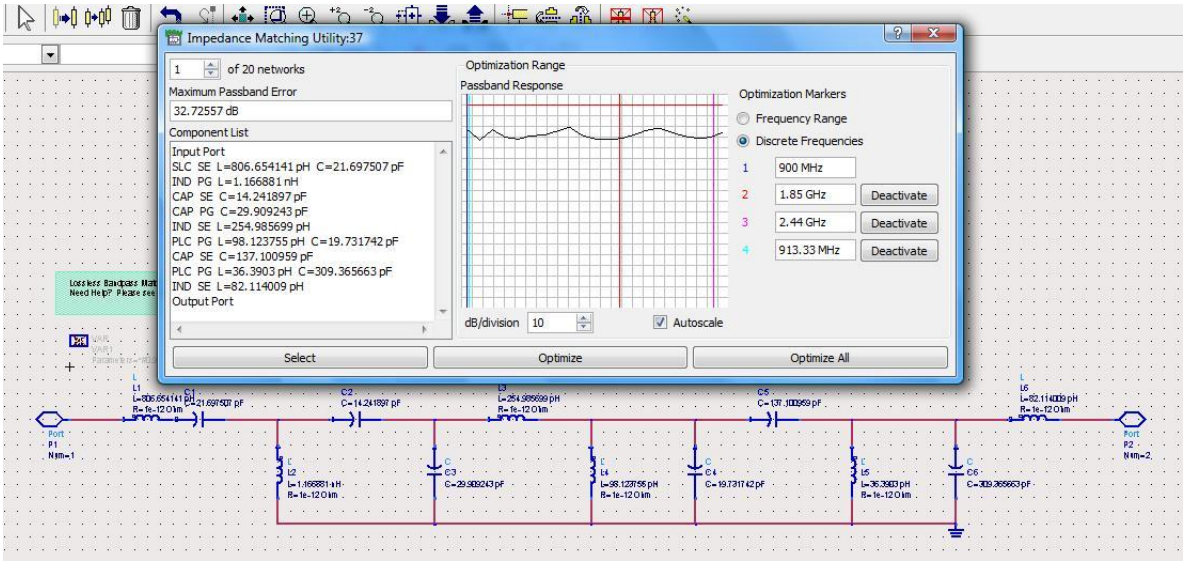


Fig. 4.3.3.4: Post Optimization for best performance in our application bands.

4.3.4 Nominal Optimization

Nominal optimization (also called Performance Optimization) is a technique to design a circuit where, under specific performance goals and a value-search method, the values of each component (or a set) are modified and then simulated, if the result of this simulation does not meet the goals, values are "searched" and simulated again until performance goals are achieved. The value-search method used for this application was the Gradient Search.

4.3.5 Gradient Search Method

The method of gradient search, adjust a set of variables according to an error function and its gradient. The simulator in the first iteration evaluates the error function and its gradient, and then the set of variables will be moved in the direction of the gradient of the error so that the error function is minimized. The simulation stops automatically when the goals are reached, when the iteration limit is reached or when there is a gradient zero. The error function used by the gradient method is the Least-Squares error function.

4.4 Simulations

Once explained the methods used for analysis and simulation of our circuits, we will show the simulations results and analysis for specific rectifier circuits. In this study we have simulated seven rectifiers that can be classified in three groups:

- Series diode rectifiers

- Parallel diode rectifier
- Two diode rectifier.

Moreover, a study of the first two groups has been made showing the differences between different methods of adaptation and different frequency bands of use.

4.4.1 Parasitic

To characterize the diode and capacitor, we took into account the parasitic. Parasitic are mechanical and electrical unwanted characteristics that limit the circuit performance. The notation used in the parasitic components is as follows

- Lp: Is an inductor placed in series with the component and can be explained as the inductance associated with the external terminations of metal that connect the internal component with the (external) circuit.
- Cp: Capacitance that is placed in parallel with the component and is due that all package solid materials have dielectric constants related that behave as capacitors.
- Rp: The metal conductor used to connect the internal component with the (external) circuit is usually a gold or gold plated with an associated very low resistance. This resistance can affect the performance of a microwave circuit.

4.4.2 Series diode rectifier matched with the ADS2008 matching utility and focusing on 900 MHz, 1.85 GHz and 2.45 GHz bands.

After optimizing the values of each component in Fig. (4.4.2.1) to achieve the maximum performance, we have obtained the values shown in Table 4.4.2.1.

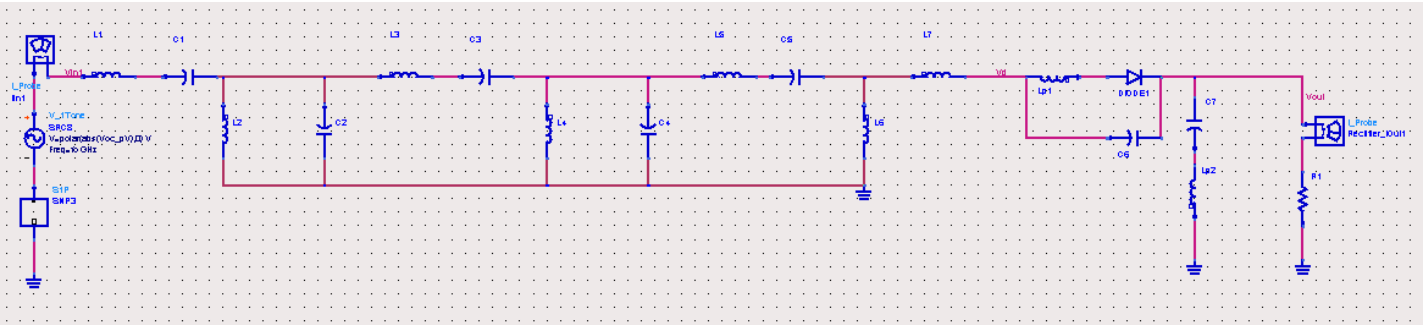


Fig. 4.4.2.1: Complete schematic: antenna S parameters, matching circuit and series diode rectifier.

Fig. 4.4.2.2 shows that the maximum efficiency for the 900 MHz, 1.85 GHz and 2.45 GHz bands is achieved when the received power is -6 dBm, -4 dBm and -6dBm respectively; we can also conclude that the maximum efficiency is 52%, 62% and 59% respectively. If we

observe the efficiency for low input power (-20 dBm), we can note that the values are 25%, 30% and 28% respectively. The main disadvantage of this circuit is the large number of components (18), if we would implement it, the soldering loss of each component would affect the performance significantly.

Table 4.4.2: Simulation values for Fig.4.4.2

Component Name	Value	Component Name	Value
L1	3.3 nH	**C1**	2.2 pF
L2	6.8 nH	**C2**	1.8 pF
L3	16 nH	**C3**	1 pF
L4	12 nH	**C4**	1.2 pF
L5	15 nH	**C5**	0.5 pF
L6	36 nH	**C6**	68 pF
L7	15 nH	**Cp1**	0.25 pF
Lp1	0.6 nH	**R1**	2.2 KOhm
Lp2	0.36 nH		

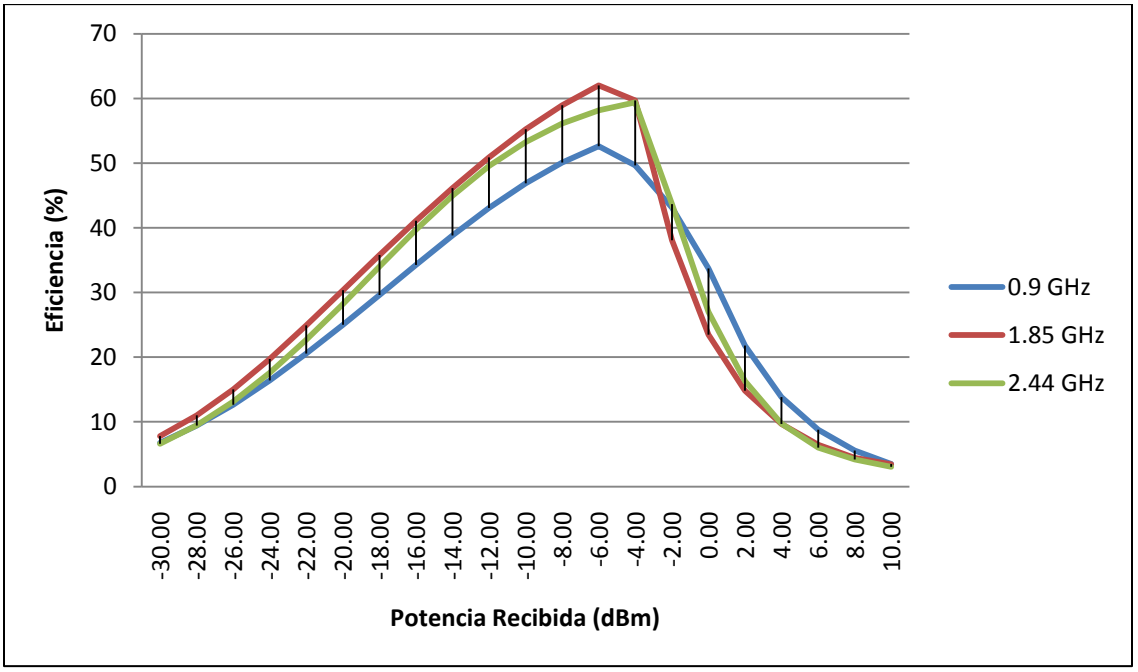


Fig. 4.4.2.2: Rectenna efficiency versus input power showing the behavior for low input power (Agilent ADS).

4.4.3 Series diode rectifier matched with LC circuit over the entire band.

After optimizing all the values of each component in Fig. (4.4.3.1) to achieve the maximum performance, we have obtained the values shown in Table 4.4.3.1.

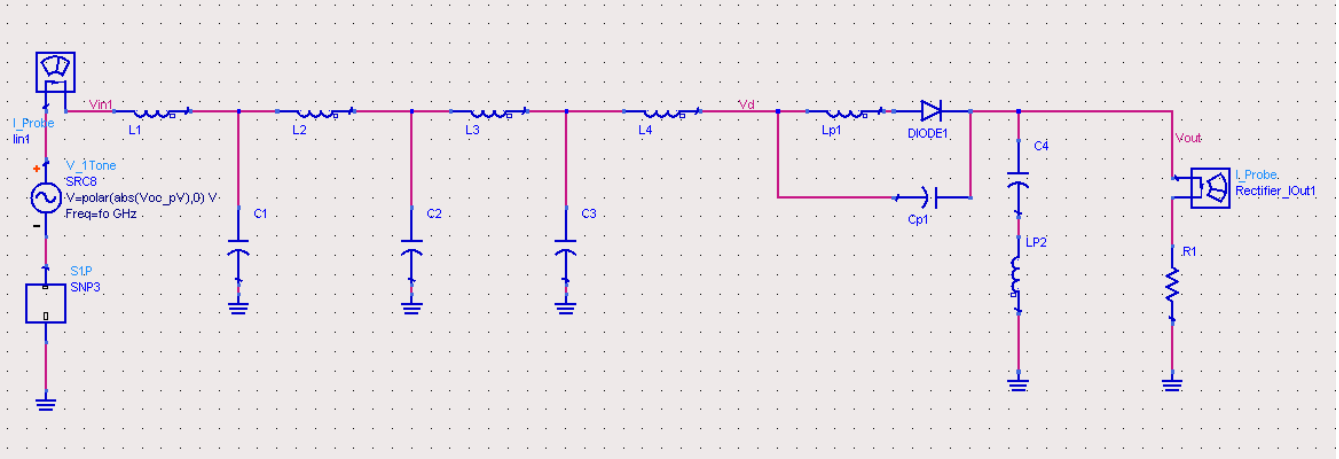


Fig. 4.4.3.1: Complete schematic: antenna S parameters, matching circuit and series diode rectifier.

Fig 4.4.3.2 shows that the maximum efficiency for the 900 MHz, 1.85 GHz and 2.45 GHz bands is achieved when the received power is -2 dBm. We can also conclude that the maximum efficiency is 44%, 41% and 50% respectively. If we observe the efficiency for low input power (-20 dBm), we can note that the values are 15.5%, 8.9% and 11% respectively, well below the Fig. 4.4.2.1 rectifier. The drop in efficiency has been remarkable, and this is because the matching over a broad band is more complicated.

Table 4.4.3: Simulation values for Fig. 4.4.3.1

Component Name	Value	Component Name	Value
L1	4 nH	**C1**	1 pF
L2	1 nH	**C2**	1 pF
L3	14 nH	**C3**	1 pF
L4	30 nH	**C4**	140 pF
Lp1	0.3 nH	**Cp1**	0.25 pF
Lp2	0.36 nH	**R1**	2.2 KOh

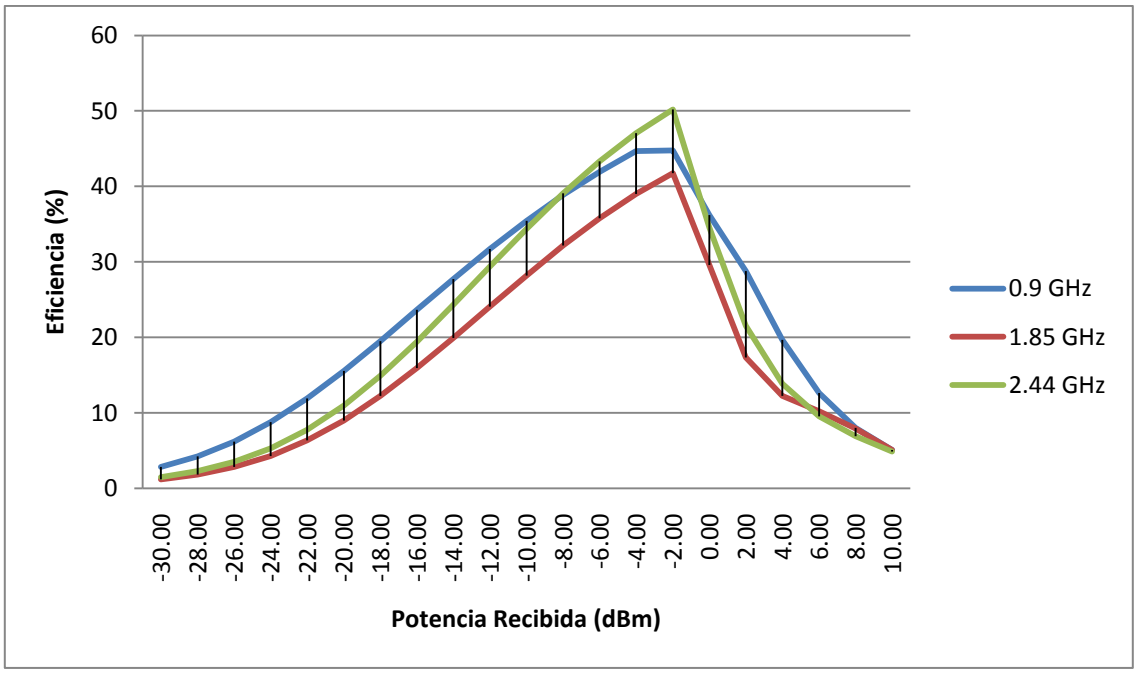


Fig. 4.4.3.2: Rectenna efficiency versus input power for low input power (Agilent ADS) at 900 MHz, 1.85 GHz and 2.45 GHz.

4.4.4 Series diode rectifier matched with LC circuit focusing on the bands of use.

After optimizing the values of each component in Fig. (4.4.4.1) to achieve the maximum performance, we have obtained the values shown in Table 4.4.4.

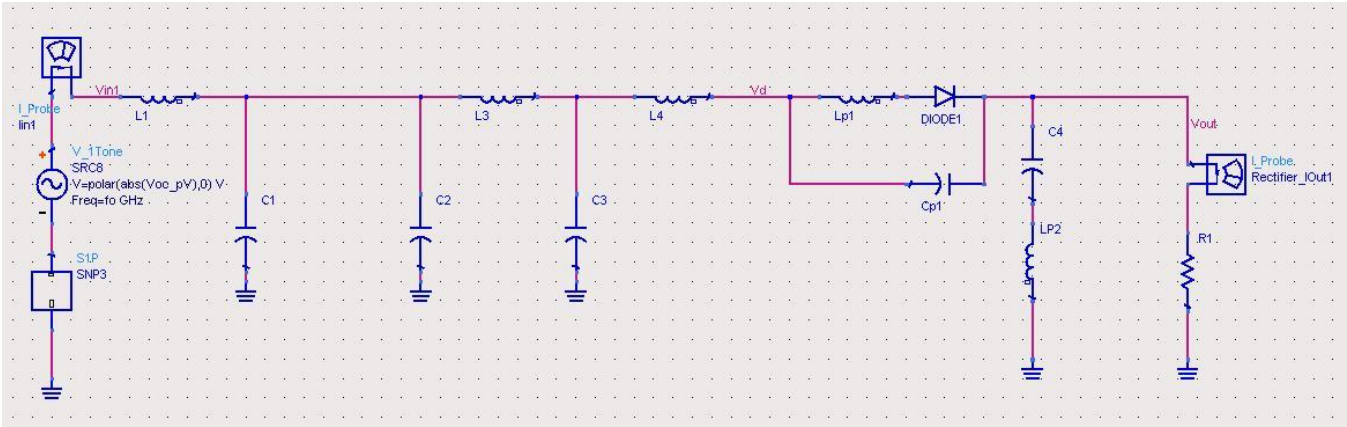


Fig. 4.4.4.1: Complete schematic: antenna S parameters, matching circuit and series diode rectifier.

Fig. 4.4.4.1 shows that the maximum efficiency for the 900 MHz, 1.85 GHz and 2.45 GHz bands is achieved when the received power is -4 dBm, -2dBm, -4dBm respectively. We can also see that the maximum efficiency is 49.2%, 44.9% and 59.8% respectively, higher than the last network. If we observe the efficiency for low input power (-20 dBm), we can note that the values are 18.7%, 13.3% and 20.9% respectively. This circuit would be a good

balance between efficiency and component amount, as the circuit of Fig. 4.4.2.1 has many components but has a very good performance and the circuit of Fig. 4.4.3.1 has few components and low performance as the band is very extensive. However, in this circuit, we have the advantage that it has fewer components than the other two shown above and also you get an acceptable efficiency.

Table 4.4.4: Simulation values for Fig. 4.4.4.1

Component Name	Value	Component Name	Value
L1	8.9 nH	**C1**	1 pF
L3	20 nH	**C2**	1 pF
L4	31.5 nH	**C3**	1 pF
Lp1	0.6 nH	**C4**	60 pF
Lp2	0.36 nH	**Cp1**	0.25 pF
		R1	2.1 KOhm

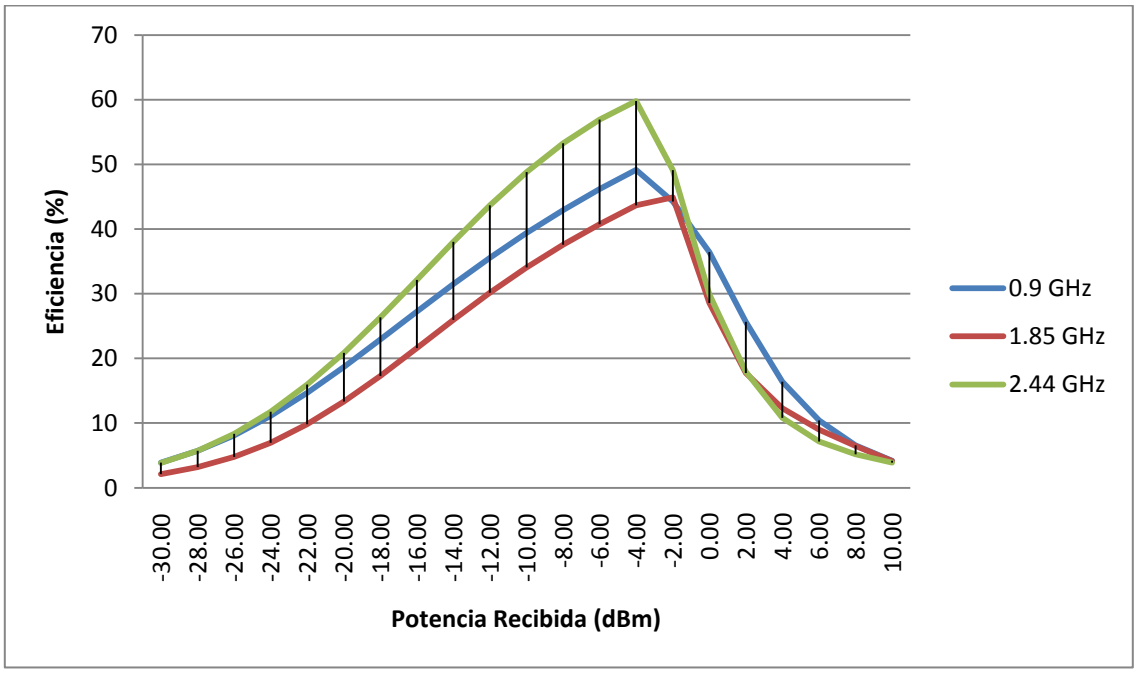


Fig. 4.4.4.2: Rectenna efficiency versus input power for low input power (Agilent ADS) at 900 MHz, 1.85 GHz and 2.45 GHz.

4.4.5 Parallel diode rectifier Matched using the ADS2008 matching utility and focusing on the bands of use.

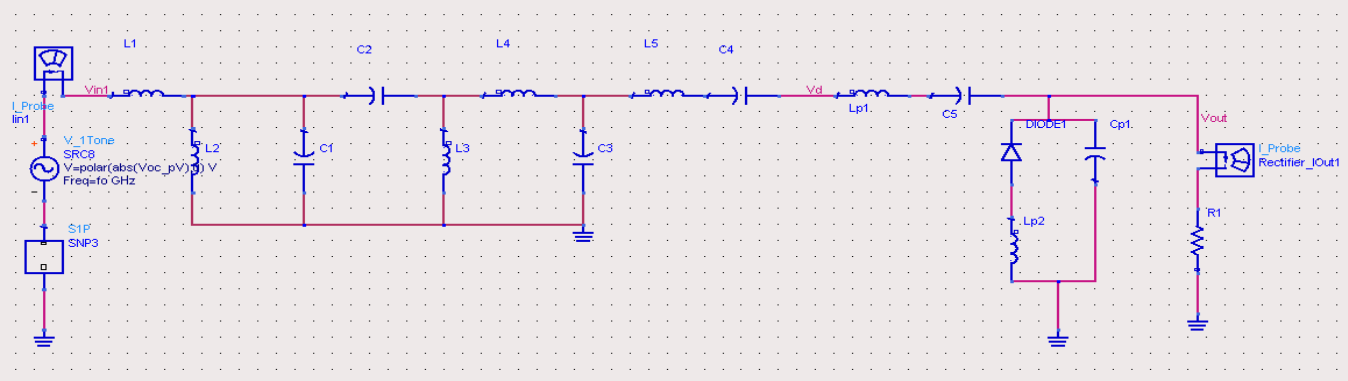


Fig. 4.4.5.1: Complete schematic: antenna S parameters, matching circuit and parallel diode rectifier.

Table 4.4.5 Simulation values for Fig. 4.4.5.1

Component Name	Value	Component Name	Value
L1	1 nH	**C1**	8.2 pF
L2	19 nH	**C2**	2 pF
L3	9 nH	**C3**	1 pF
L4	15 nH	**C4**	300 pF
L5	24 nH	**C5**	112 pF
Lp1	0.6 nH	**Cp1**	0.25 pF
Lp2	0.36 nH	**R1**	5.4KOh

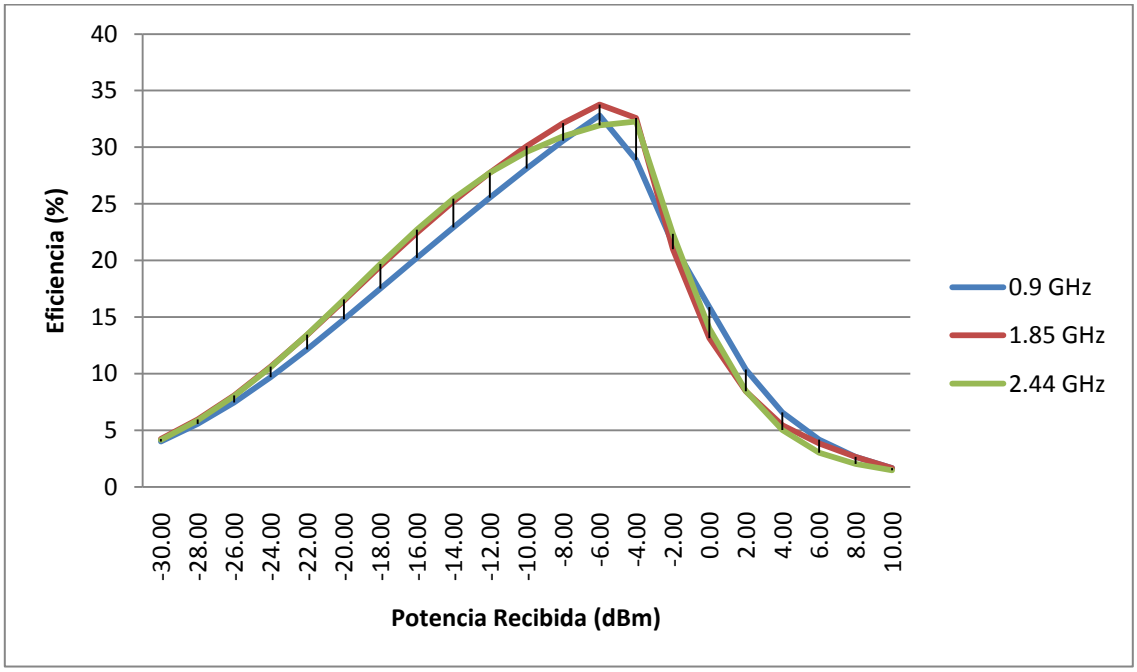


Fig. 4.4.5.2: Rectenna efficiency versus input power for low input power (Agilent ADS) at 900 MHz, 1.85 GHz and 2.45 GHz.

4.4.6 Parallel diode rectifier matched with LC circuit over the entire band.

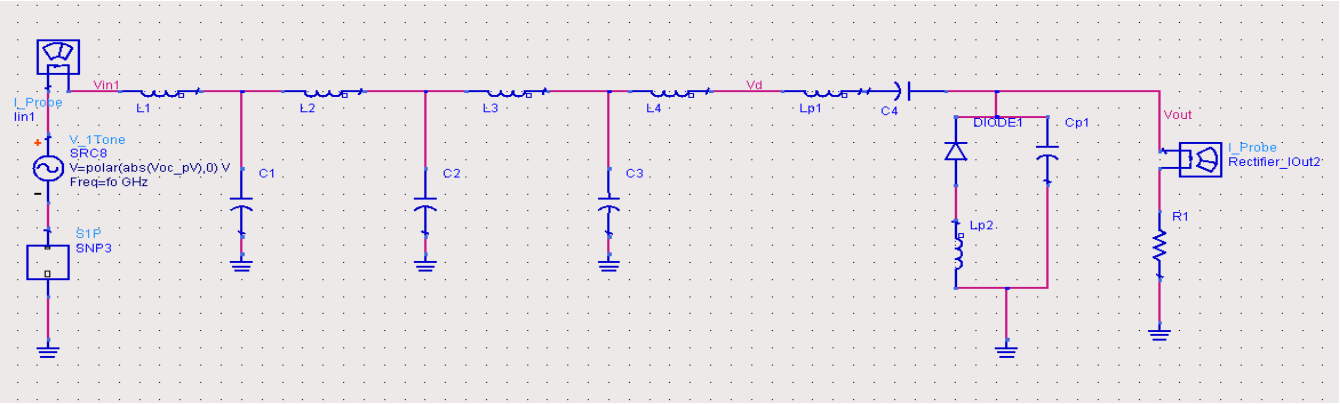


Fig. 4.4.6.1: Complete schematic: antenna S parameters, matching circuit and parallel diode rectifier.

Table 4.4.6: Simulation values for Fig. 4.4.6.1

Component Name	Value	Component Name	Value
L1	4 nH	**C1**	1.5 pF
L2	8.2 nH	**C2**	1 pF
L3	8 nH	**C3**	1 pF
L4	25 nH	**C4**	60 pF
Lp1	0.6 nH	**Cp1**	0.25 pF
Lp2	0.36 nH	**R1**	3.5KOh

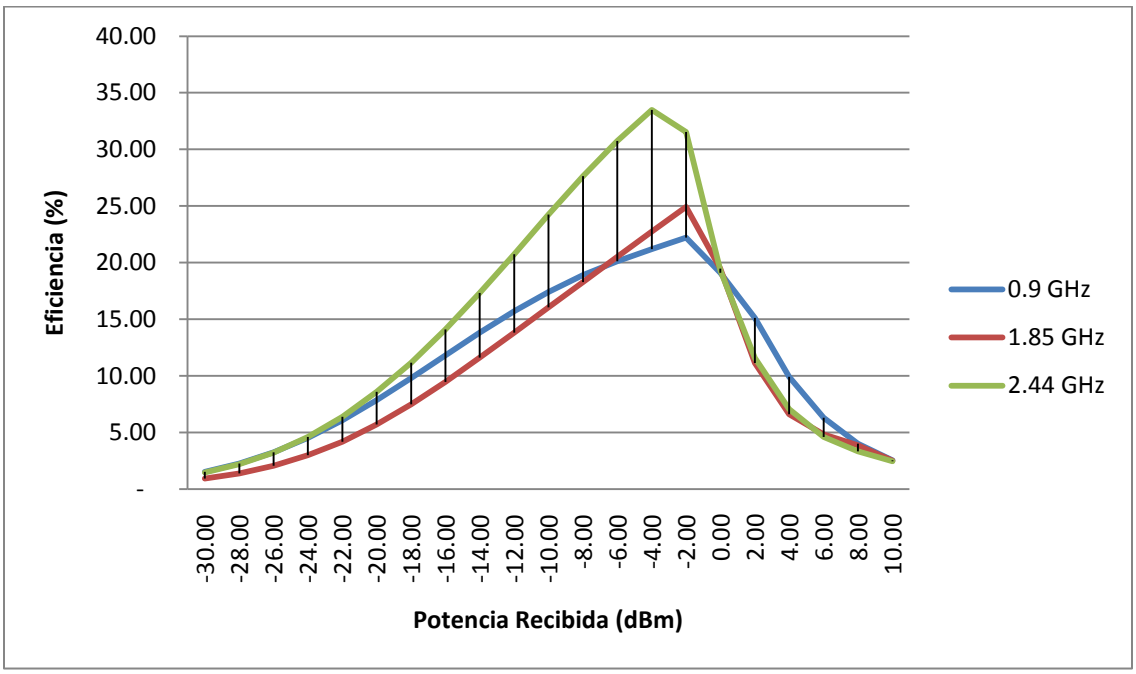


Fig. 4.4.6.2: Rectenna efficiency versus input power for low input power (Agilent ADS) at 900 MHz, 1.85 GHz and 2.45 GHz.

4.4.7 Parallel diode rectifier matched with LC circuit focusing on the bands of use.

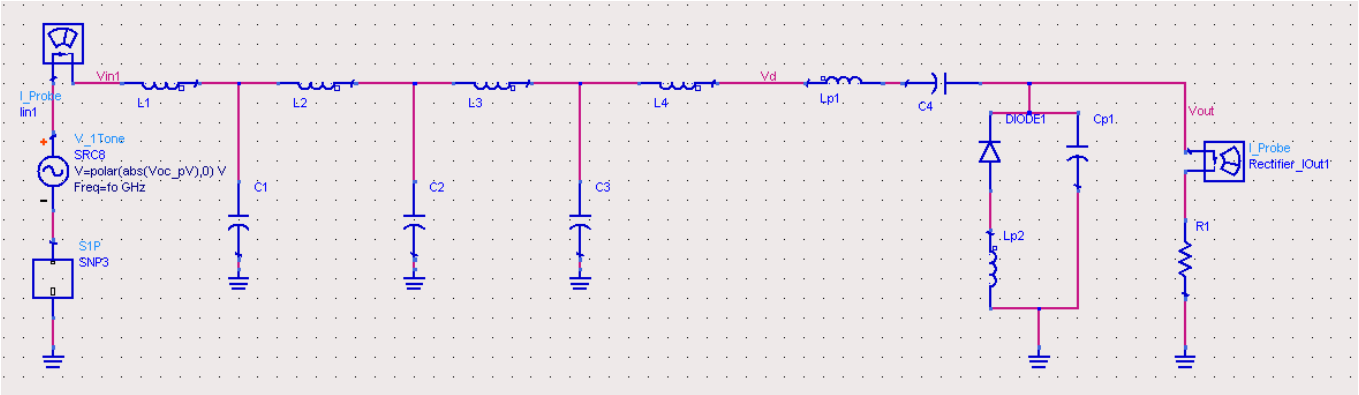


Fig. 4.4.7.1: Complete schematic: antenna S parameters, matching circuit and parallel diode rectifier.

Table 4.4.7: Simulation values for Fig. 4.4.7.1

Component Name	Value	Component Name	Value
L1	2.2 nH	**C1**	2 pF
L2	11.6 nH	**C2**	1 pF
L3	24 nH	**C3**	1 pF
L4	32 nH	**C4**	90 pF
Lp1	0.6 nH	**Cp1**	0.25 pF
Lp2	0.36 nH	**R1**	4.3KOh

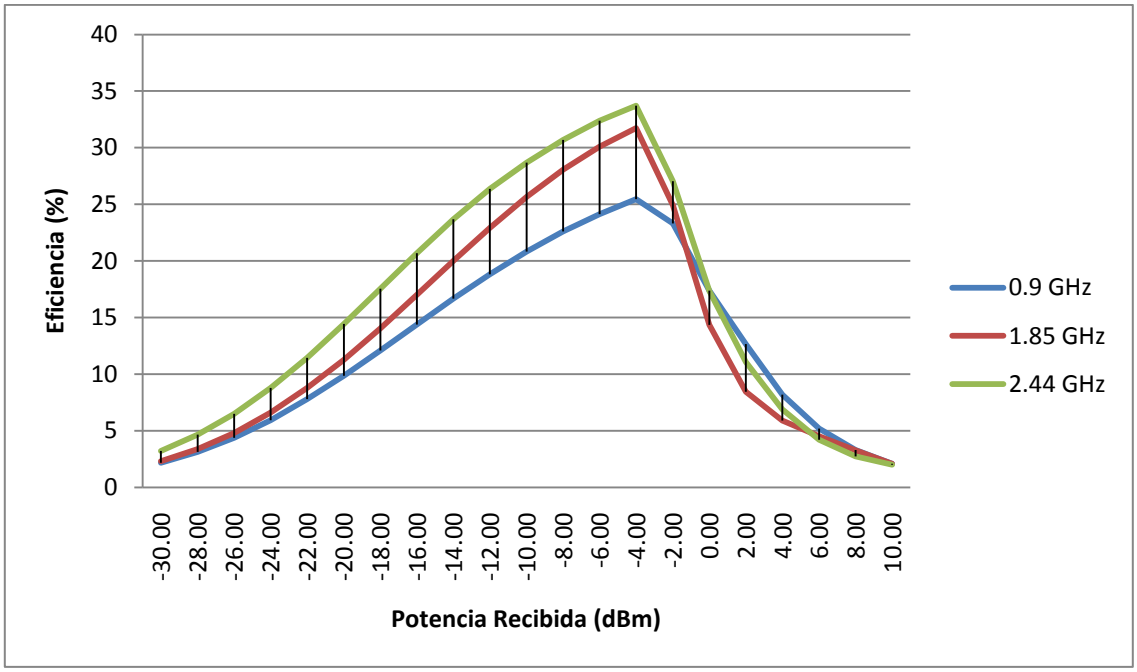


Fig. 4.4.7.2: Rectenna efficiency versus input power for low input power (Agilent ADS) at 900 MHz, 1.85 GHz and 2.45 GHz.

4.4.8 Two diode rectifier matched using the ADS2008 Impedance Matching Utility and focusing on the bands of use.

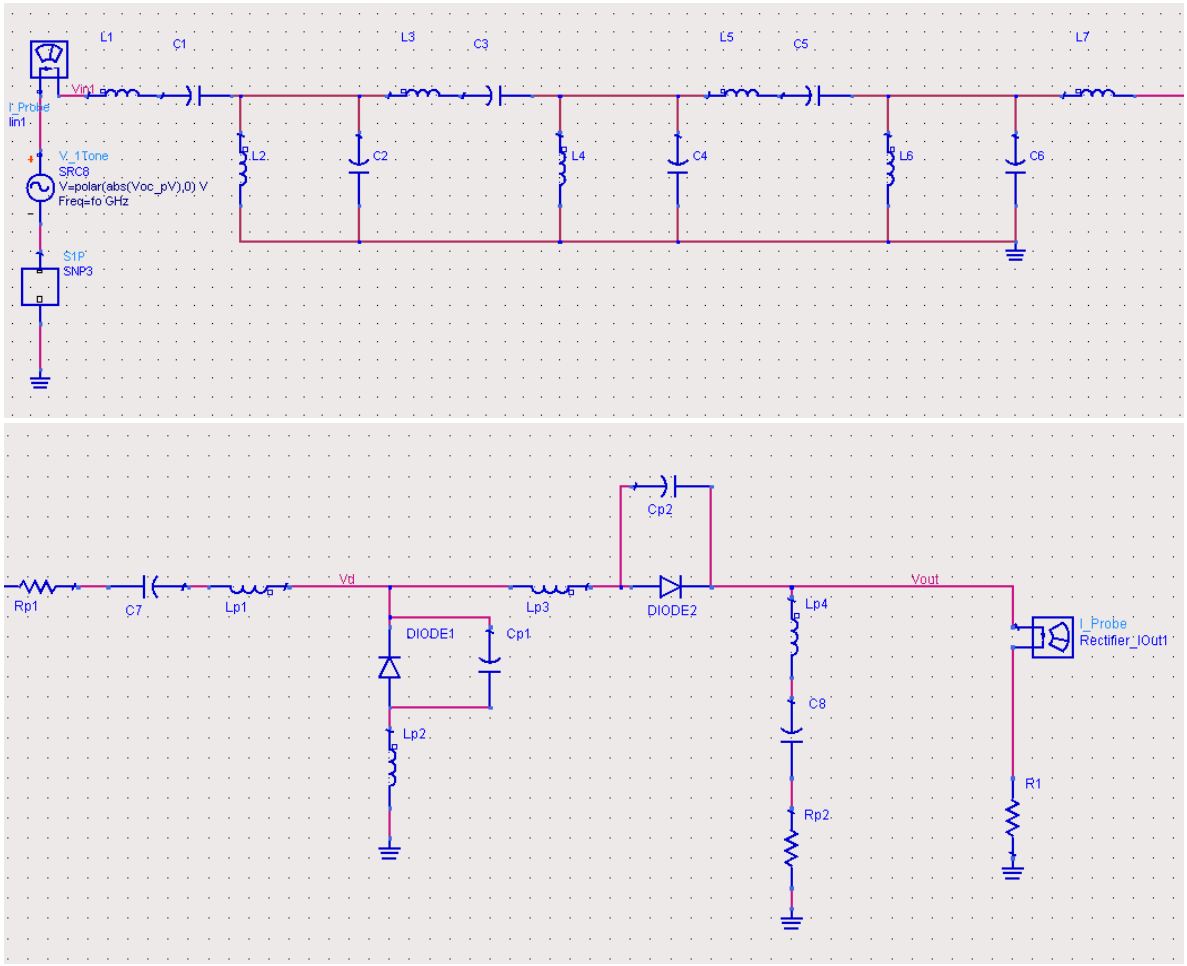


Fig. 4.4.8.1: Complete schematic: antenna S parameters, matching circuit and two diode rectifier.

Table 4.4.8: Simulation values for Fig. 4.4.8.1

Component Name	Value	Component Name	Value
L1	9.1 nH	**C1**	1.5 pF
L2	6.4 nH	**C2**	2.6 pF
L3	6.2 nH	**C3**	2.5 pF
L4	6.5 nH	**C4**	4 pF
L5	13 nH	**C5**	0.7 pF
L6	8.8 nH	**C6**	2.2 pF
L7	9.6 nH	**C7**	63 pF
Lp1	0.6 nH	**Cp1**	0.25 pF
Lp2	0.36 nH	**Cp2**	0.25 pF
Lp3	0.6 nH	**R1**	6.6KOh
Lp4	0.36 nH		

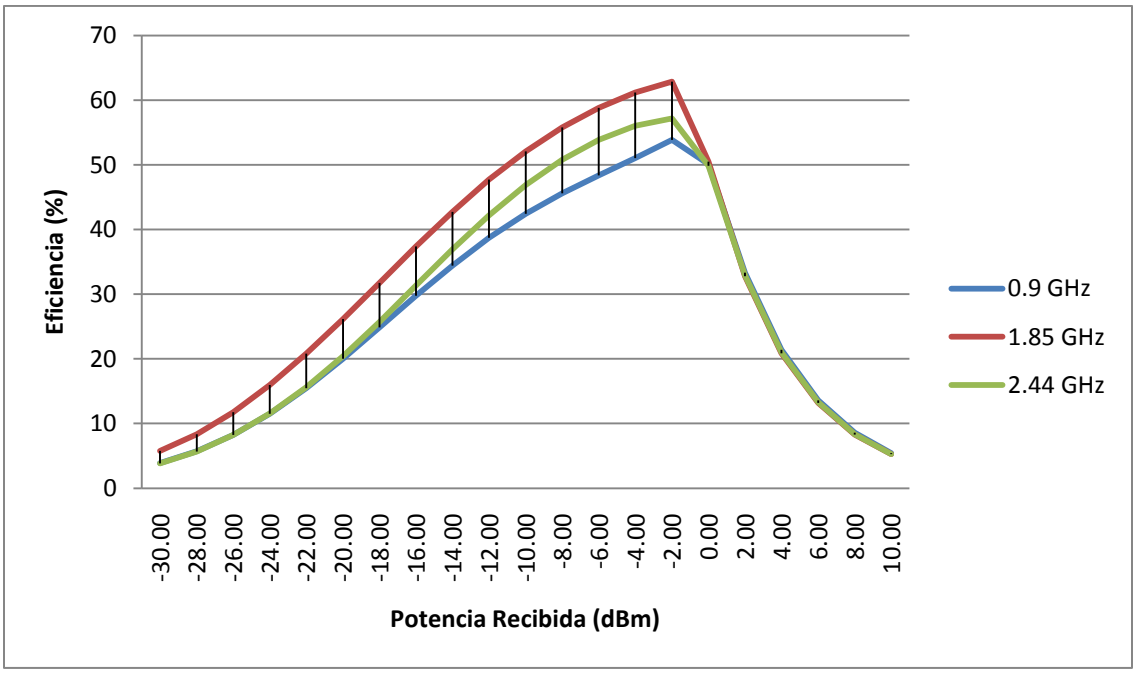


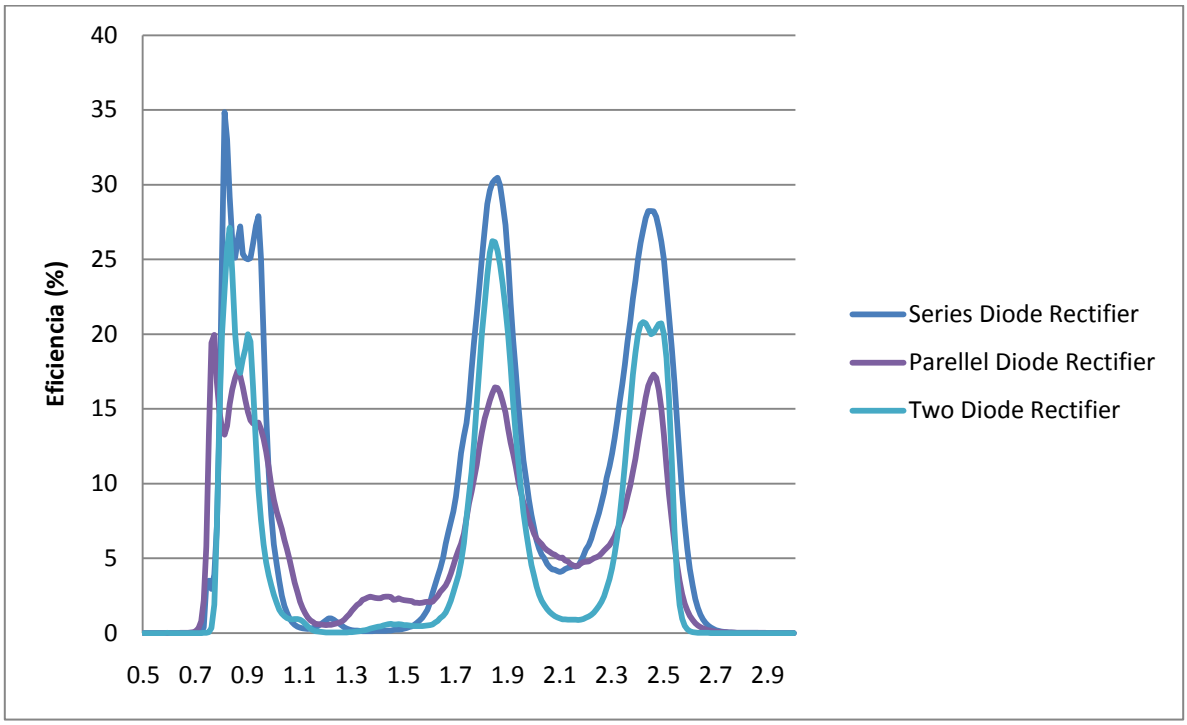
Fig. 4.4.8.2: Rectenna efficiency versus input power for low input power (Agilent ADS) at 900 MHz, 1.85 GHz and 2.45 GHz.

4.5 Efficiency Analysis for Low Input-power.

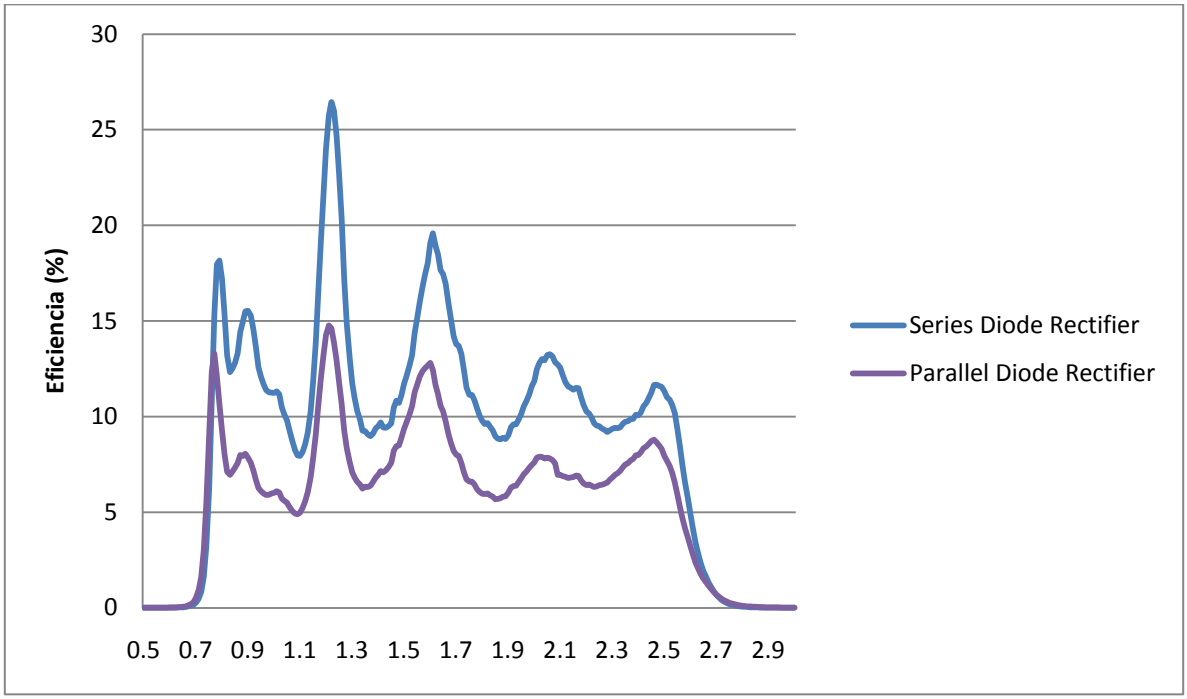
After analyzing the efficiency of each band in a power range, we will now analyze the efficiency when the received power is -20 dBm (which is a good example of an electromagnetic signal in the environment) and for frequencies ranging from 500 MHz to 3 GHz.

Fig. 4.5.1 (a) shows the efficiency of the rectifiers that have been adapted using the ADS2008 matching utility. A better performance is appreciated in the series rectifier; the parallel diode rectifier, shows an efficiency about 13% less for all bands while the two diode rectifier, in the 900 MHz, 1.85 GHz and 2.45 GHz bands has an efficiency of 8%, 5% and 8% respectively, lower than the series diode rectifier. As explained above, the difficulty with this method of adaptation is the great quantity of components required and the consequent loss of performance when soldering. Fig. 4.5.1 (b) shows the efficiency versus frequency for the rectifiers adapted with LC circuits and for all 700 MHz- 2.5 GHz band, we can observe that the results are similar but less efficient than other methods of adaptation, improved efficiency is achieved with the rectifier diode series that has reached about 5% higher efficiency than the parallel rectifier diode; this low efficiency is the result of adapting a broad band of frequencies with a limited number of components. Fig. 4.5.2 shows the efficiency of the rectifier adapted with LC circuit over the bands of interest, again we see that a higher performance was obtained with the series diode rectifier diode,

this is because it is more feasible to adapt rectifier using the same number of components to smaller frequency bands.



(a)



(b)

Fig. 4.5.1: Efficiency versus frequency for -20dBm input power for (a) the rectifiers adapted with the ADS2008 matching utility and with focus over the 900 MHz, 1.85 GHz and 2.45 GHz bands (b) the rectifiers adapted with LC circuits for the entire 700 MHz – 2.5 GHz band.

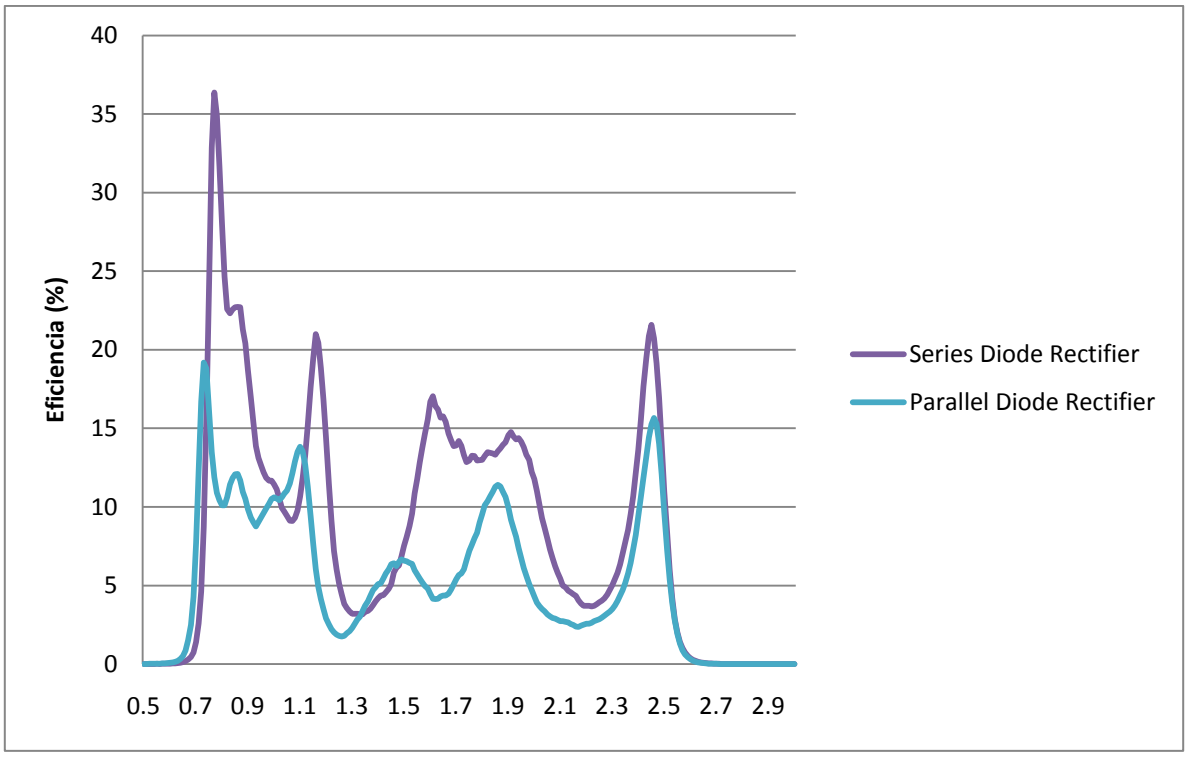


Fig. 4.5.2: Efficiency versus frequency for -20dBm input power for the rectifiers adapted with LC circuit with focus over the 900 MHz, 1.85 GHz and 2.45 GHz bands

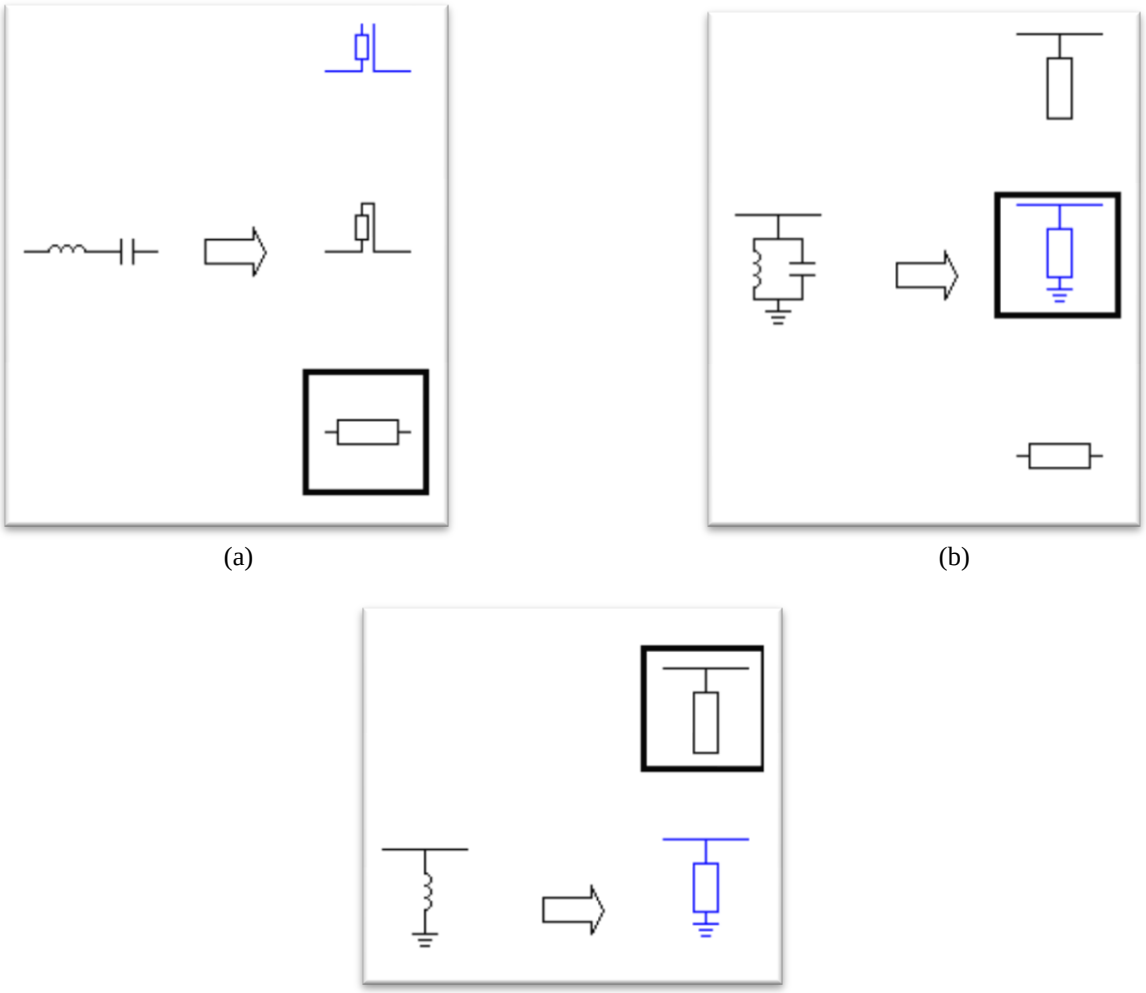
4.6 LC to Microstrip-Lines Transformation

According to the simulations shown above, the series diode rectifier was chosen to proceed with the application for showing the best performance in simulations; the problem now would focus on choosing the proper matching method. Given the low efficiency of the rectifier circuit adapted with LC elements, we have decided to improve the rectifier adapted with the ADS2008 matching utility but with a LC-Microstrip line transformation to avoid the effects of soldering and to exploit the characteristics of microstrip lines.

Fig. 4.6.1 shows the equivalence for the transformation from LC elements to microstrip lines; we can see that the series inductor and capacitor have been transformed into a transmission line, the parallel grounded inductor and capacitor have been transformed into a grounded parallel stub and the parallel grounded inductor has been transformed into a not grounded parallel stub. Fig. 4.6.1 shows the complete circuit schematic where the inductors and capacitors have been transformed into transmission lines.

To improve the efficiency of our circuit, the techniques of harmonic balance and nominal optimization were applied again. In this case, the parameters that were optimized were: the lengths and widths of each transmission line, the inductor value, the rectifier capacitor and the load resistance. Table 4.6 shows the final values of optimization. Fig. 4.6.2 shows the

efficiency as a function of the frequency when the input power is -20 dBm, a good example of electromagnetic signal in the environment.



(a) (b) (c)

Fig. 4.6.1: (a) series inductor and capacitor (b) parallel grounded inductor and capacitor (c) parallel grounded inductor possible transformations, components in a box were selected

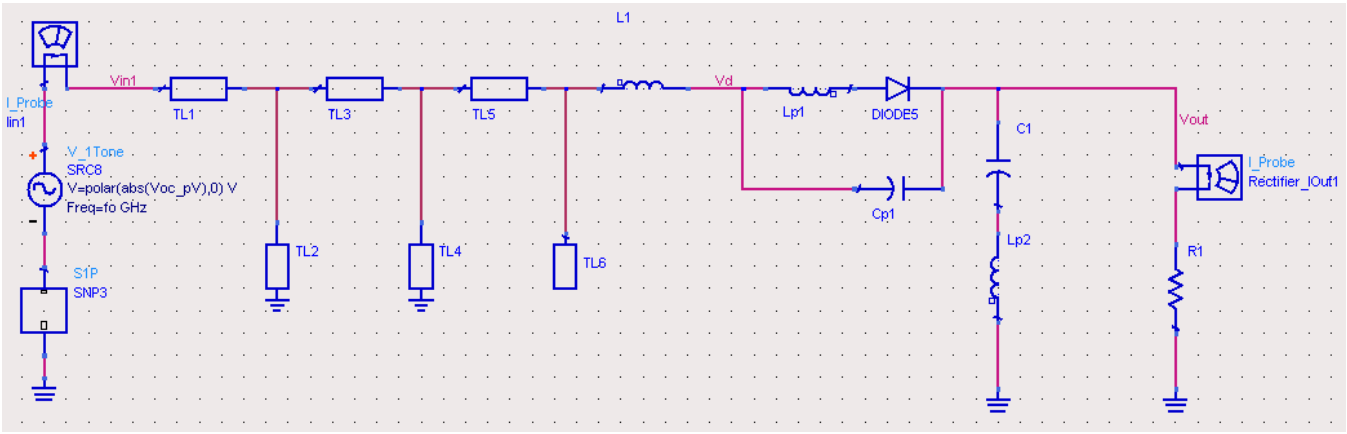


Fig. 4.6.2: Complete schematic: antenna S parameters, matching microstrip lines and the series diode rectifier.

The Fig. 4.6.3 also shows a comparison of efficiency as a function of frequency between the classical LC circuit components and the new transmission lines circuit. We can observe that the efficiency is almost identical with the exception of some small details such as that the implementation of transmission lines has improved by 5% the 900 MHz band performance and the 1.85 GHz band shows efficiency drop of 4%.
In Fig. 4.6.4 a comparison of the efficiency as a function of received power shown, we can observe that the maximum efficiency for 900 MHz, 1.85 GHz and 2.45 GHz is achieved when the received power is -6 dBm, - 6 dBm and -4 dBm respectively and the efficiency at this values are 48.6%, 64.8% and 53.6% respectively.

Table 4.6: Simulation values for Fig. 4.6.1

Component Name	Value	Component	Value
L1	12 nH	**TL5**.L	2.7 cm
TL1.L	3.2 cm	**TL5**.A	0.3 mm
TL1.A	3.5 mm	**TL6**.L	0.25 cm
TL2.L	2.7 cm	**TL6**.A	1.8 mm
TL2.A	0.4 mm	**C1**	100 pF
TL3.L	2.7 cm	**Cp1**	0.25 pF
TL3.A	0.8 mm	**Lp1**	0.36 nH
TL4.L	2.4 cm	**Lp2**	0.6 nH
TL4.A	0.73 mm	**R1**	0.36 nH

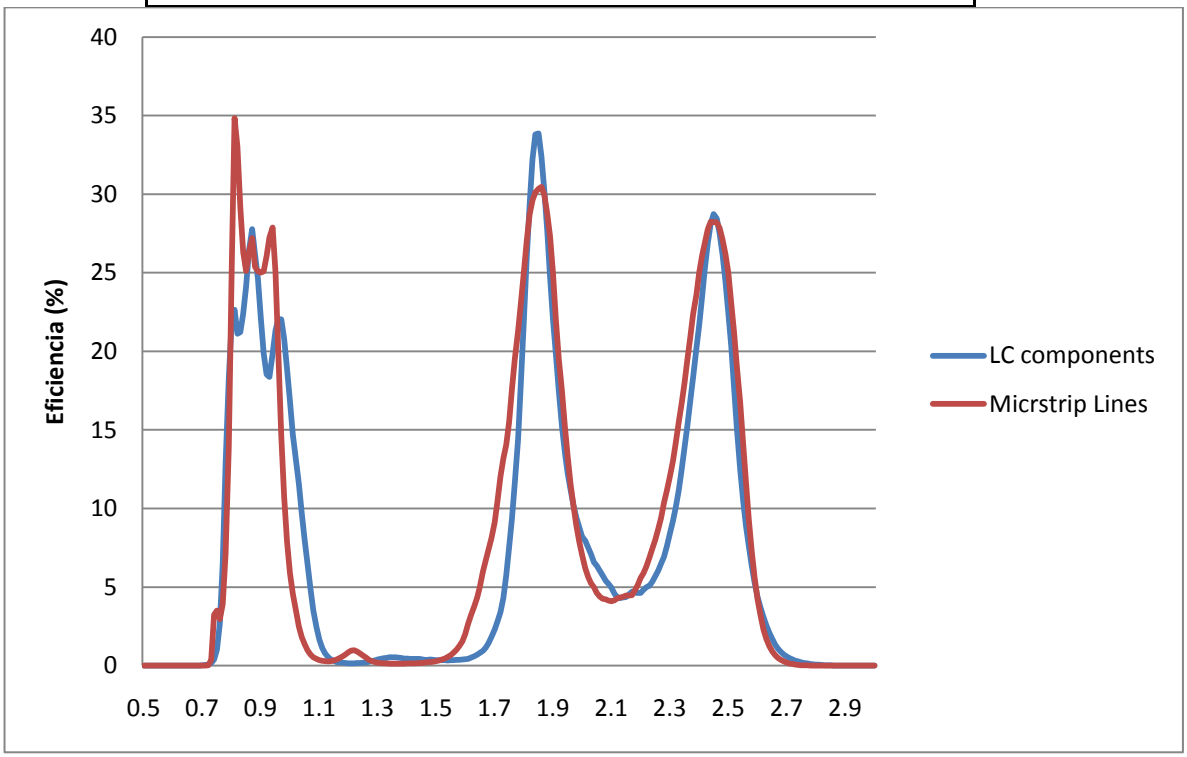


Fig. 4.6.3: Efficiency versus frequency for -20dBm input power for the rectifiers adapted with LC components and Microstrip Lines for the 900 MHz, 1.85 GHz and 2.45 GHz bands

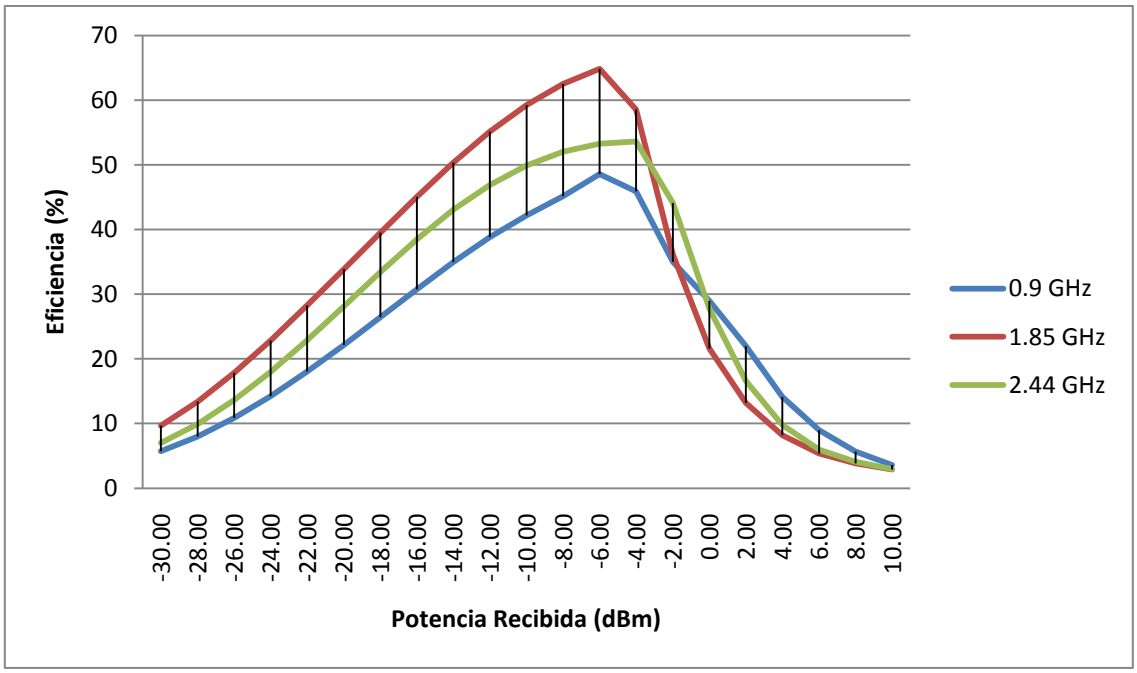


Fig. 4.6.4: Rectenna efficiency versus input power for low input power (Agilent ADS) at 900 MHz, 1.85 GHz and 2.45 GHz.

4.6.1 Momentum

Momentum is an electromagnetic simulator that computes S-parameters for various topologies such as ours that presents vias (holes that connect two layers of the topology). This tool is very useful to predict circuit performance at high frequencies and one of the main advantages of momentum is that, from a schematic, you can create a physical layout to simulate the characteristics of the substrate and also allocate the necessary ports to simulate the different components of our circuit. Momentum is based on the numerical discretization method called method of moments.

In general, the method of moments, fragments transmission lines into smaller segments to find the current distribution in each of these, so at the end of the simulation, we can know the joint behavior of these to characterize the entire system.

4.6.2 Momentum Layout

Once we have the LC component values optimized as well the transmission lines lengths, the next step is to generate the layout of our matching circuit to simulate the transmission lines together with the substrate. In Fig. 4.6.2.1 the auto generated matching circuit layout is shown.

According to the values in Table 4.6, the lengths of transmission lines L1, L3 and L5, are 3.2cm, 2.7cm and 2.7cm respectively, giving a total of 8.6cm.

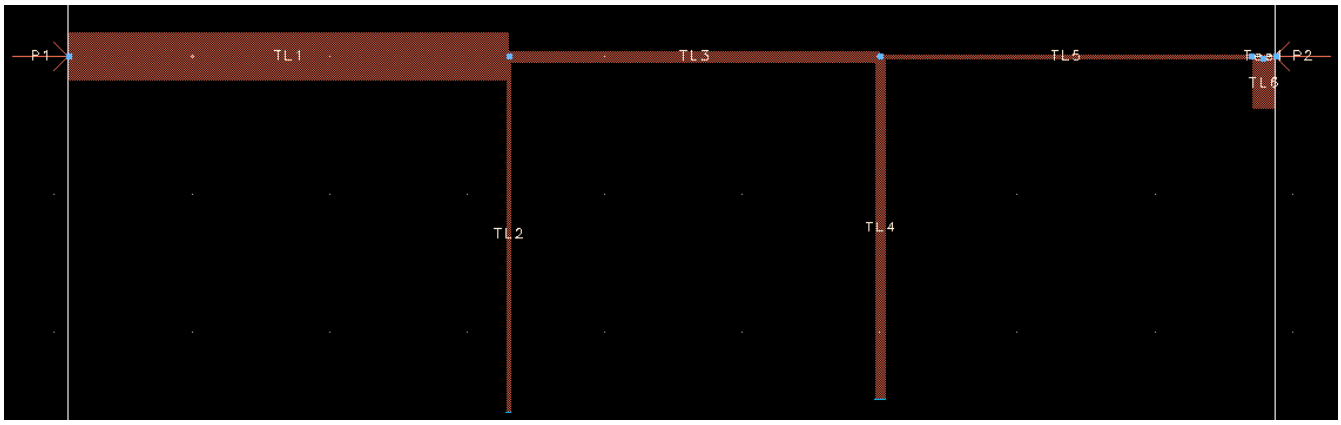


Fig. 4.6.2.1: auto generated matching circuit layout by ADS2008 Momentum

To improve the circuit size and make it more compact, a technique has been applied; we have kept the lengths of the transmission lines but instead of straight lines we have used curved transmission lines. The measures of the complete layout with straight lines were: 8.75cm x 2.75cm, while the measurements of the size-improved layout are: 2,97cm x 2.31cm, 66% and 16% shorter.

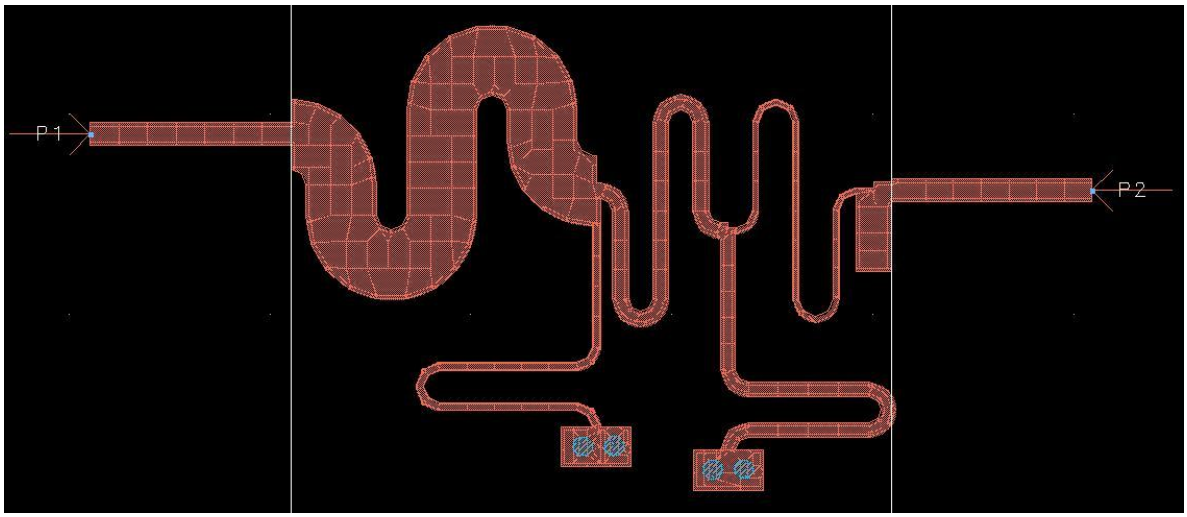


Fig. 4.6.2.2: size-improved matching circuit and momentum meshing.

In Fig. 4.6.2.1 and 4.6.2.2, we can observe that the ground lines are different, this is because the ADS2008 generates an ideal model of ground connection, which is not desirable when simulating since the simulations should adjust to real parameter, what is we really want, that's why in Fig. 4.6.2.2 the ideal ground has been replaced by a transmission line with two vias that connects the top layer with the ground plane through the

substrate. This hole will be subsequently filled with a conductor paste; this will be discussed in detail in Section 4.7.

Since the results after simulating with Momentum could be different and therefore not reach the goals, some adjustments in the transmission lines should be made; we have seen empirically that by adjusting the length of the third stub, the 2.45 GHz band moves to the right or the left if ,respectively, the length decreases or increases. A similar behavior has been observed in the other stubs, so that the efficiency for the frequency bands has been adjusted successfully. Fig. 4.6.2.3 shows the efficiency as a function of frequency for the rectifier with the data that Momentum has generated. To implement momentum data in the simulation of the entire circuit, the implementation of a data module is necessary. Fig. 4.6.2.4 shows the module has been implemented.

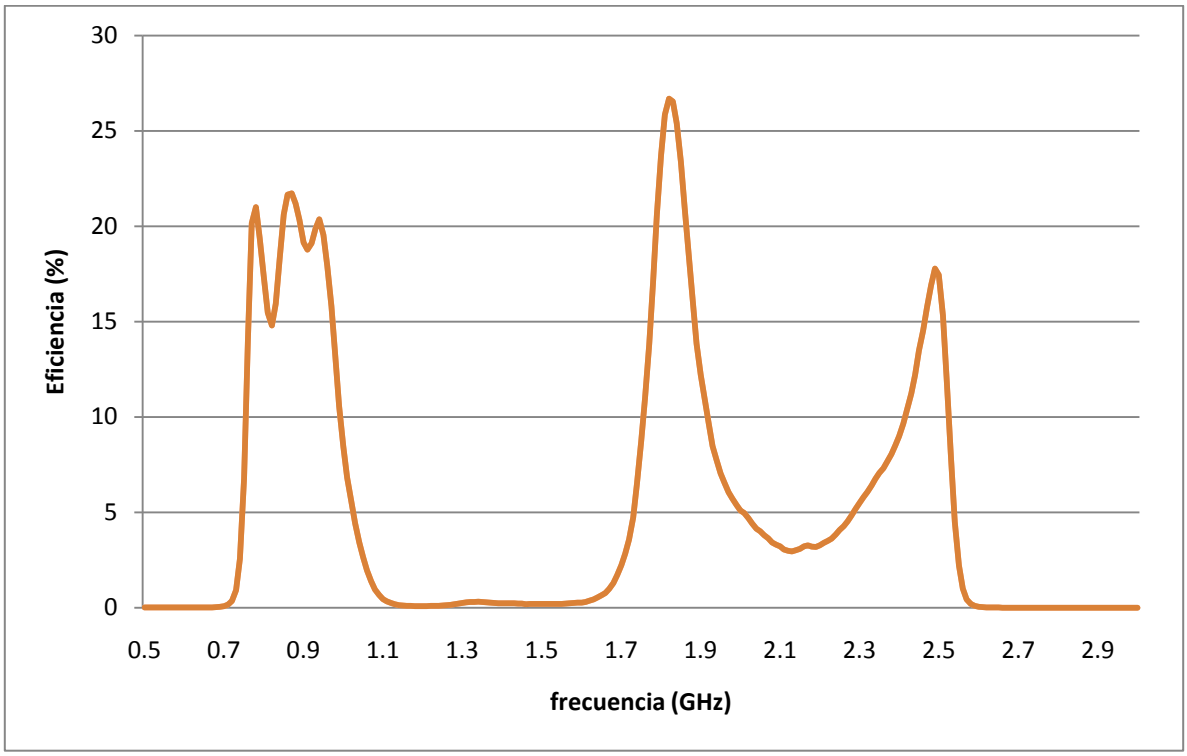


Fig. 4.6.2.3: Efficiency versus frequency showing the behavior for -20dBm input power with the matching circuit Momentum data implemented.

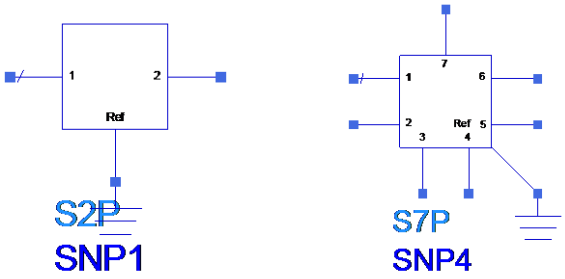


Fig. 4.6.2.4: 2 and 7-port S-Parameter module, the data generated by Momentum is loaded into these items and then simulated with the other circuit components.

The rectifier layout was designed based on the measurements of the components to be attached. Fig. 4.6.2.5 shows the layout of the rectifier that, for the schematic simulation, components such as diodes, inductors, capacitors and resistors are replaced by ports; each of these ports corresponds to one pin of the module data (Fig. 4.6.2.4) in the schema.

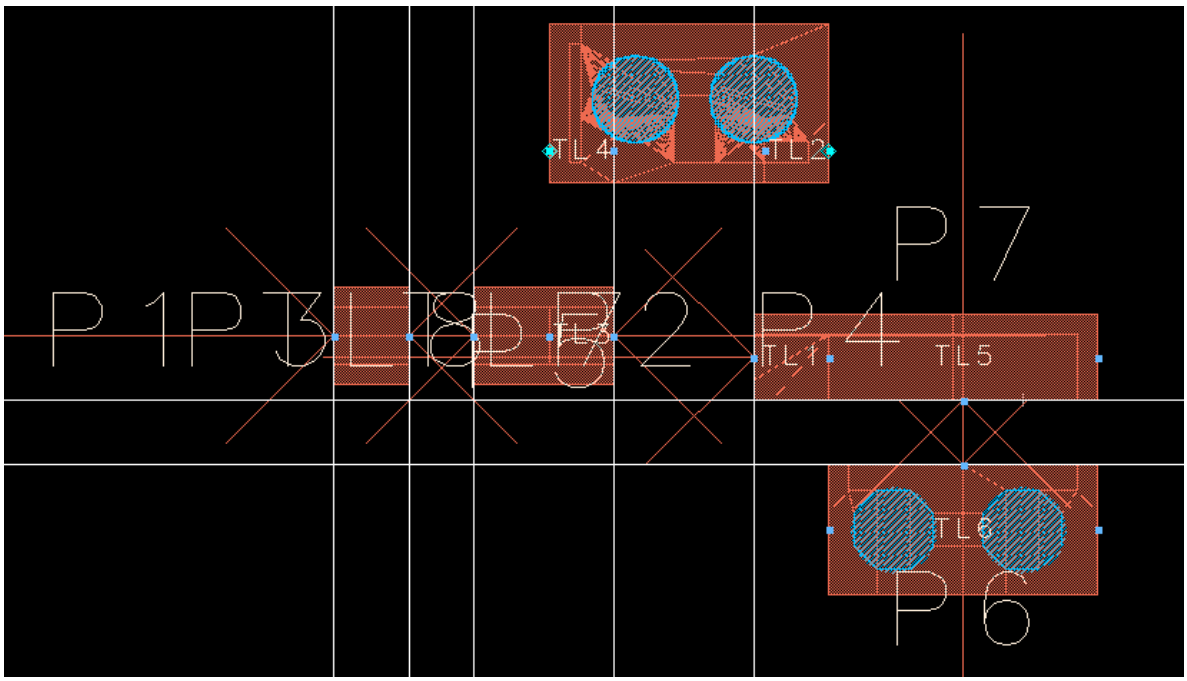


Fig. 4.6.2.5: Rectifier layout, the dimensions are 7.1 x 5.3mm

Fig. 4.6.2.6 shows the complete schematic with the integrated data modules. The first module, has the matching circuit momentum simulation data, the second module, has the momentum simulation data of the transmission lines of the rectifier, so that the ports that had been placed before in the layout now are in the respective pins on the module to simulate with the components.

Fig. 4.6.2.7 shows the efficiency versus frequency, we can observe that the maximum efficiency for the 900 MHz, 1.85 GHz and 2.45 GHz bands is about 20%. Fig. 4.6.2.8 shows the efficiency for the power received, we can see that for the bands mentioned above, the maximum efficiency is 48.6%, 64.8% and 53.6% respectively and are achieved when the received power is of -6 dBm,-4dBm and -6dBm respectively.

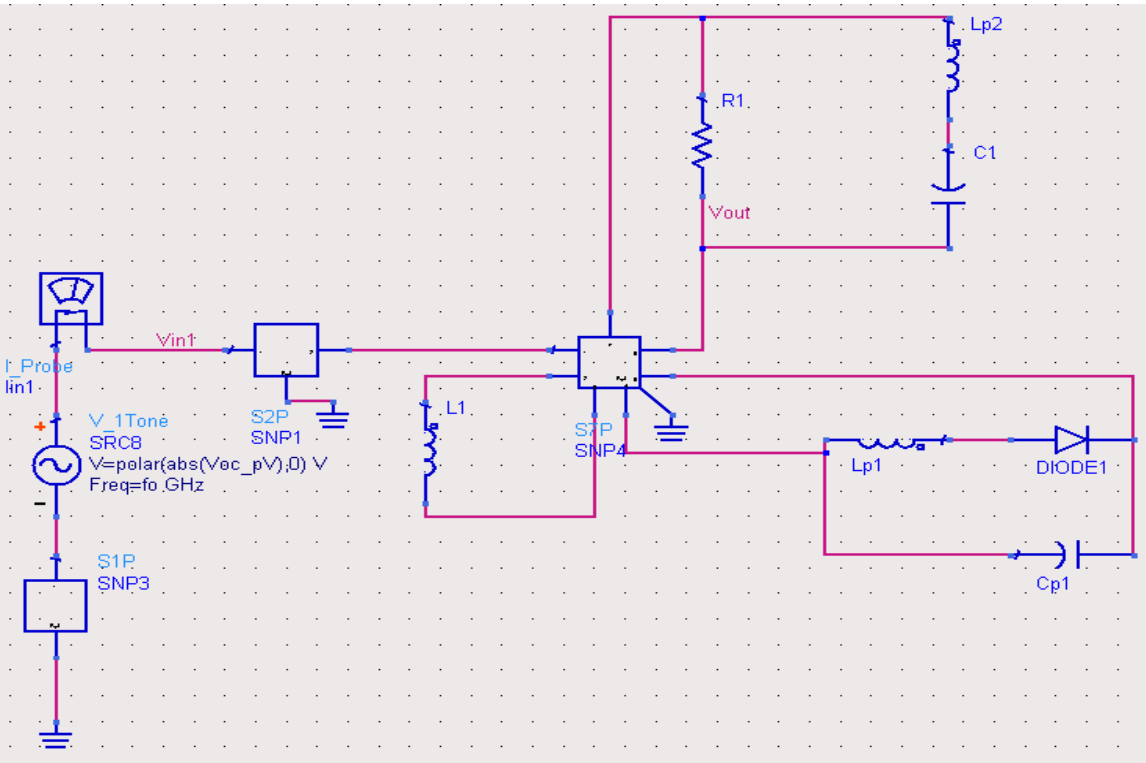


Fig. 4.6.2.6: Complete final schematic including matching lines and rectifier transmission lines momentum data

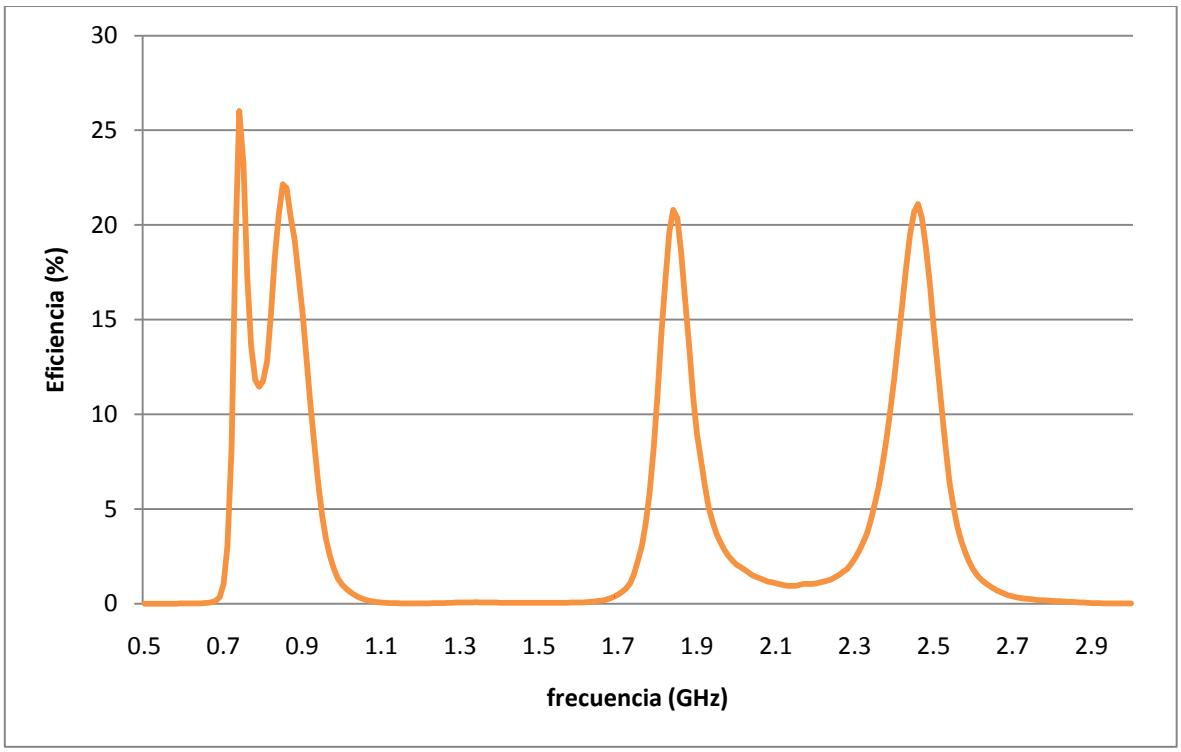


Fig. 4.6.2.7: Efficiency versus frequency showing the behavior for -20dBm input power with the matching and rectifier lines Momentum data implemented.

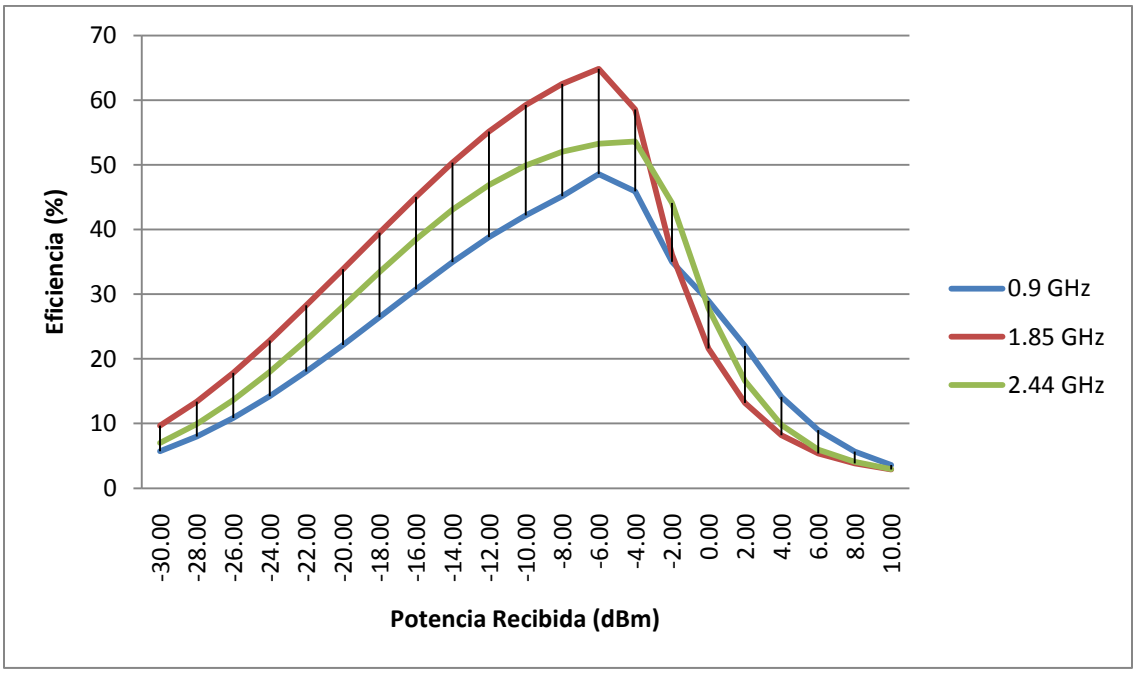


Fig. 4.6.2.8: Rectenna efficiency versus input power for low input power (Agilent ADS) at 900 MHz, 1.85 GHz and 2.45 GHz.

4.7 Rectifier Fabrication and Measurements

The manufacturing process of the rectifier circuit, as it is made from the same substrate as the antenna, is almost the same. The only difference is that to make the VIA holes, we need to follow a procedure called metallization of circuit boards [21]. The first step of this procedure is to protect the circuit with a plastic protective film before drilling the holes. The second step is the application of the LPKF Pro Conduct ® paste that, with the help of a vacuum table to suck it, will coat the walls of the hole. Finally, after removing the plastic protective film, the circuit is introduced in a hot air oven at 160 º C for 30 minutes.

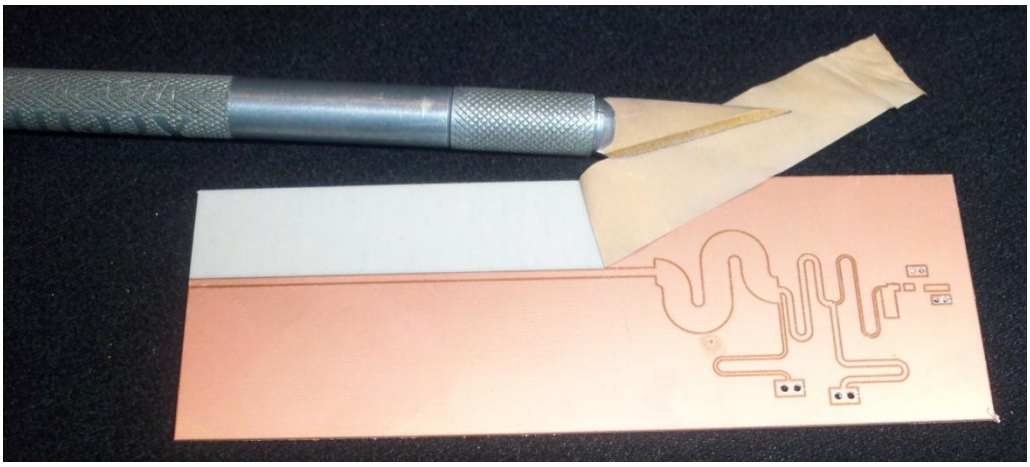
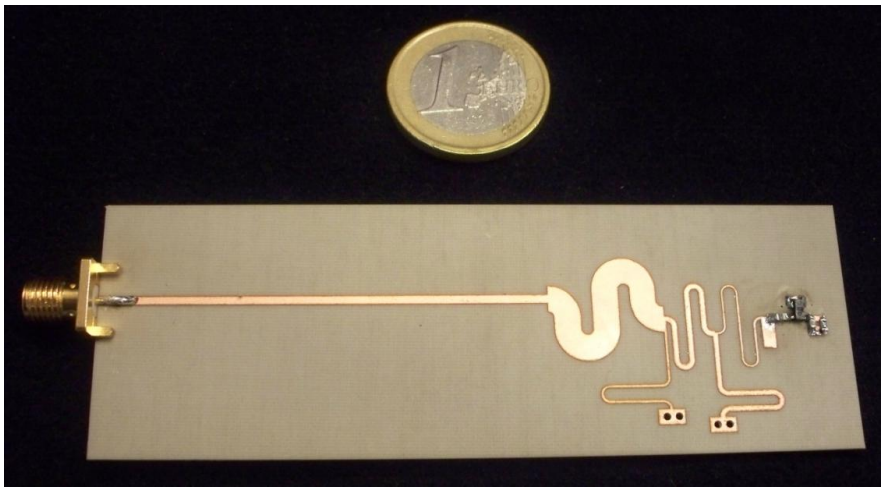
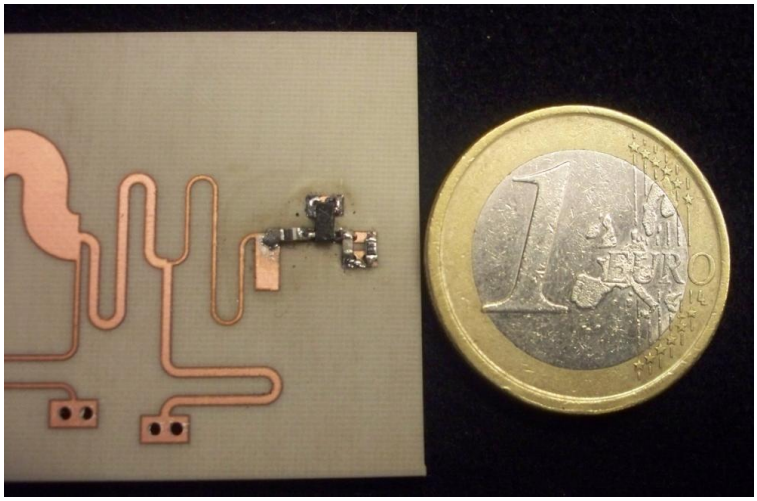


Fig. 4.7.1: Recent fabricated matching-rectifier circuit with copper excess.



(a)



(b)

Fig. 4.7.2: shows the matching circuit (a), the rectifier circuit (b) and a zoom to the soldered components(c)

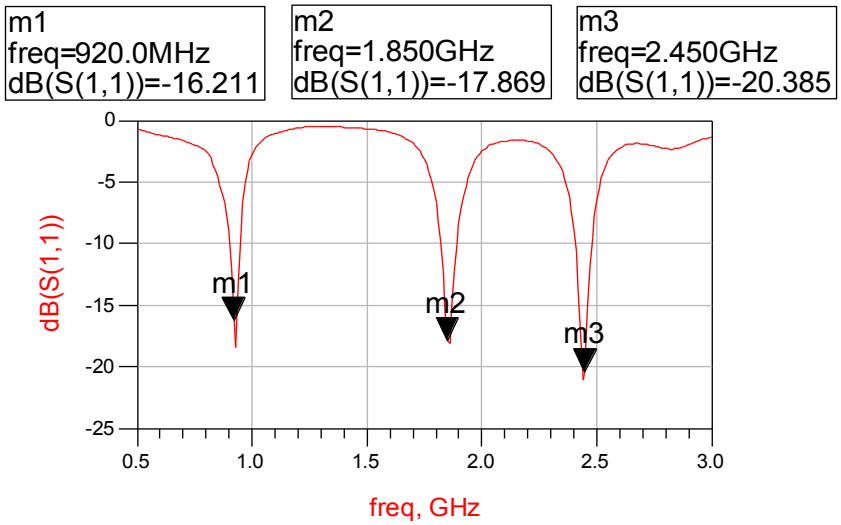


Fig. 4.7.3: shows the S11 parameters of the rectifier.

Fig. 4.7.1 shows, similarly to the antenna fabrication, the way to remove the copper excess that covers the top layer of the substrate with a cutter; and Fig. 4.7.2, as a form of demonstration, shows the dimensions of the newly manufactured circuit compared with a euro coin to the side. Note that the exact measurements have been detailed in section 4.2.6.
The parameters measured and compare with the simulations were the S-parameters (section 4.3.1). To carry out these measurements a Rohde & Schwarz ZVA24 Vector Network Analyzer (VNA) has been used.

Fig. 4.7.4: Rectenna measurement: analog signal generator and multimeter

The measurements for the full rectification circuit were made, therefore, with the matching circuit attached to it. To make these measurements, we used the Agilent E8257D analog signal generator, an Agilent E4407B spectrum analyzer and a Promax MC-859 digital multimeter. The first thing we did was to "calibrate" the cable which would connect the rectifier circuit to the analog signal generator, this is done because minimal DC power is lost in the cable; for this "calibration" the same cable to be used later was connected from the analog signal generator to the spectrum analyzer in order to obtain more accurately delivered power measures from the signal generator. After this, we connect the rectifier to the signal generator, and with the help of a multimeter, the output DC voltage was measured.

Once we have the measurement system ready, we can proceed to take some measures. The measurements were taken as follows:

- An input power sweep, ranging from -25 dBm to 5 dBm for each frequency (900 MHz, 1.85 GHz and 2.44 GHz).
- A frequency sweep, ranging from 800 MHz to 2.5 GHz for an input power of -10 dBm.
- A frequency sweep, ranging from 800 MHz to 2.5 GHz for an input power of -20 dBm.

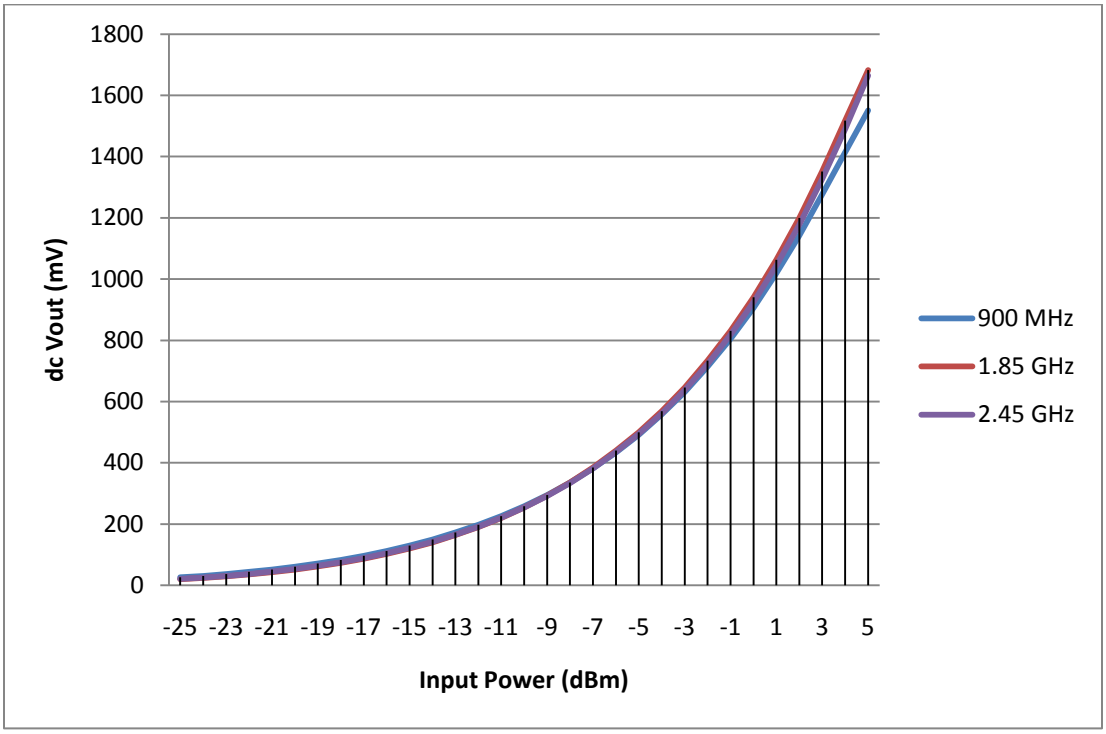


Fig. 4.7.5: Output voltage vs. input power for 900MHz, 1.85 GHz and 2.45GH

Fig. 4.7.4 shows that as we increase the input-power, the DC voltage increases proportionally, but this does not necessarily indicate that the overall efficiency shows the same behavior. This can be justified by observing the Fig. 4.7.5, where the efficiency vs. input-power plot, denotes that for high values of input power, efficiency begins to fall. For efficiency values: First, we calculated the output power by squaring DC voltage measured with the multimeter (which we have previously converted from mV to V) and was then divided by our radiation resistance. Second, we calculated the RF power by transforming the input power from dBm to Watts and finally, we divided the value of the output power by the RF power value.

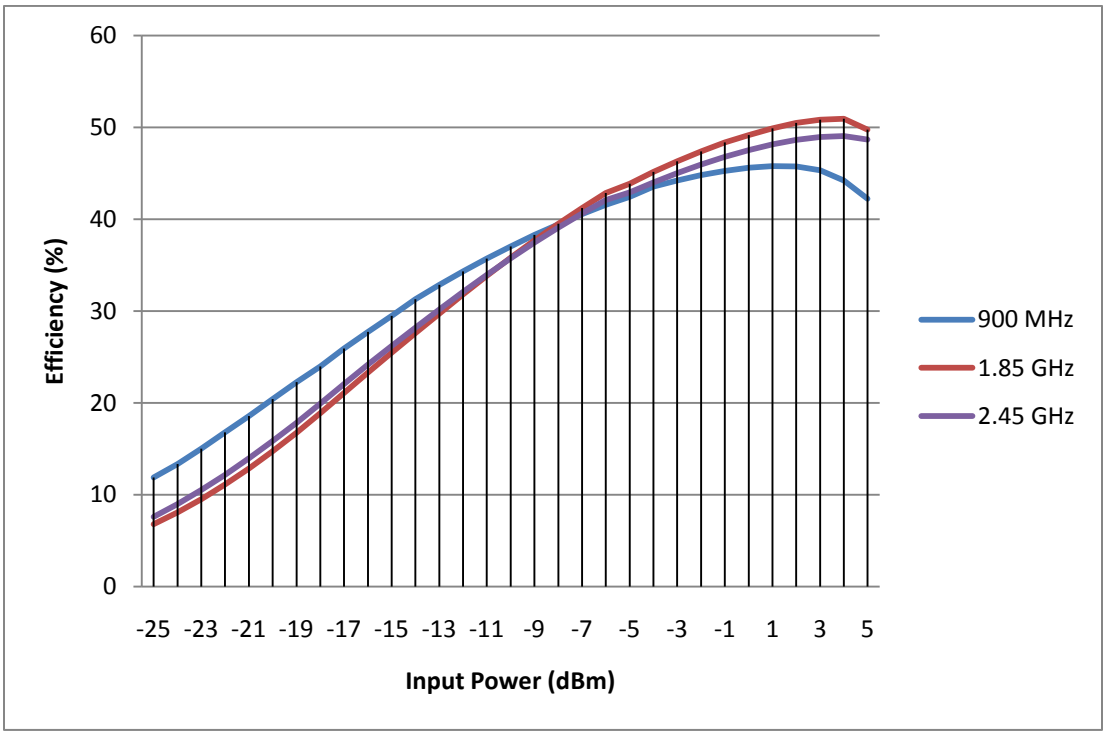


Fig. 4.7.6: Efficiency vs. input power for 900MHz, 1.85 GHz and 2.45GH

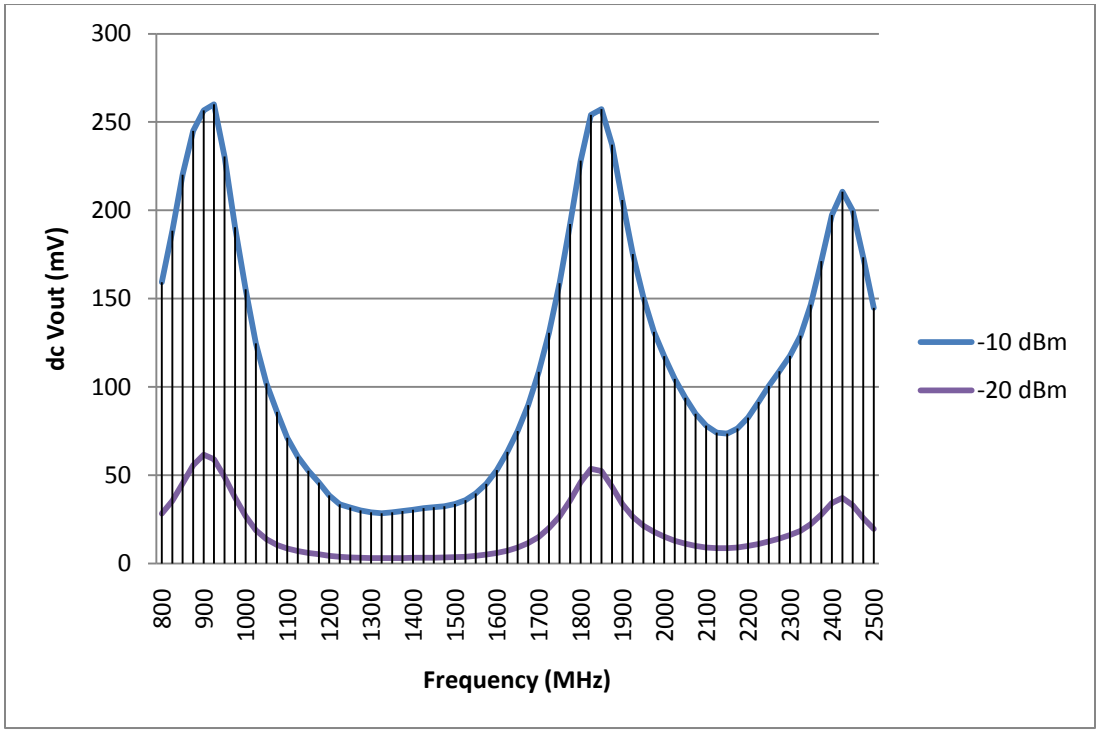


Fig. 4.7.7: Output voltage vs. frequency for -10 and -20 dBm input power

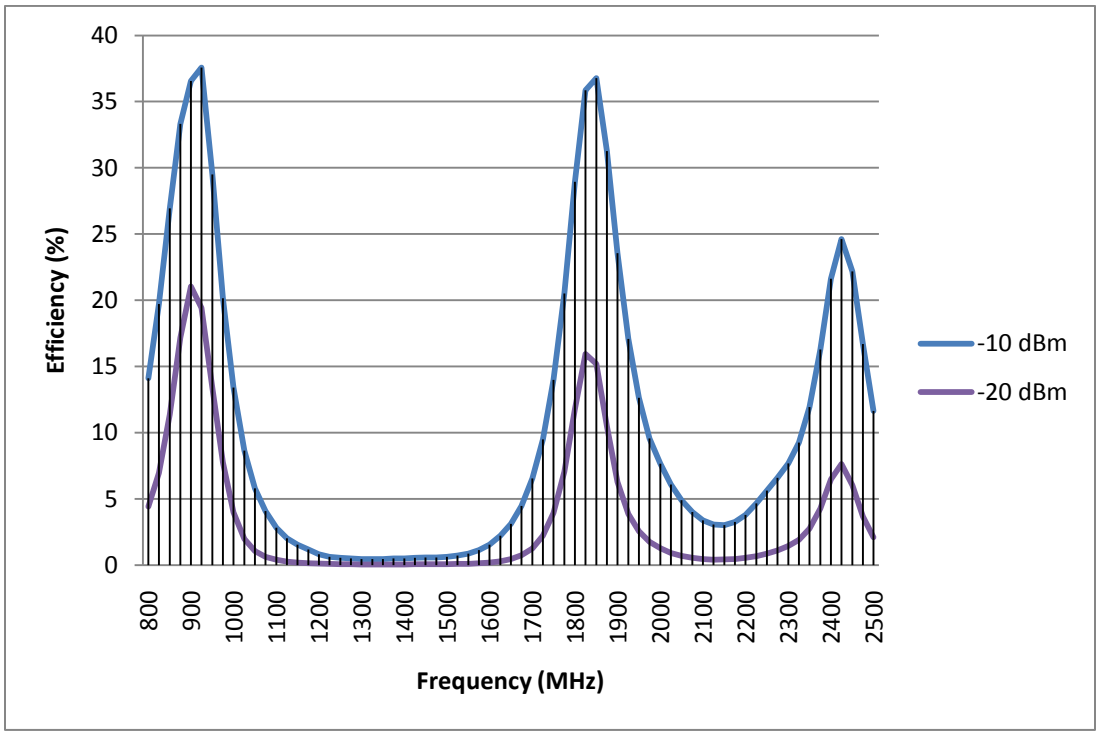


Fig. 4.7.8: Efficiency vs. frequency for -10 and -20 dBm input power

Fig. 4.7.6 shows measurements of the output voltage for a known input power, but now, making a frequency sweep ranging from 800 MHz to 2.5 GHz. The results show a better performance in the frequency bands that are involved in our application for both -10 dBm and -20 dBm input power values. We can conclude it that proper rectifier circuit design and fabrication has been done. A similar behavior is shown in Fig.4.7.7where the efficiency shows a similar pattern to the output voltage. Can be seen that for 900 MHz, 1.85 GHz and 2.45 GHz bands, the efficiency values with -10 dBm input power are around 37%, 36% and 24 % and for -20 dBm input power, the efficiency values are around 21%, 16% and 7% respectively.

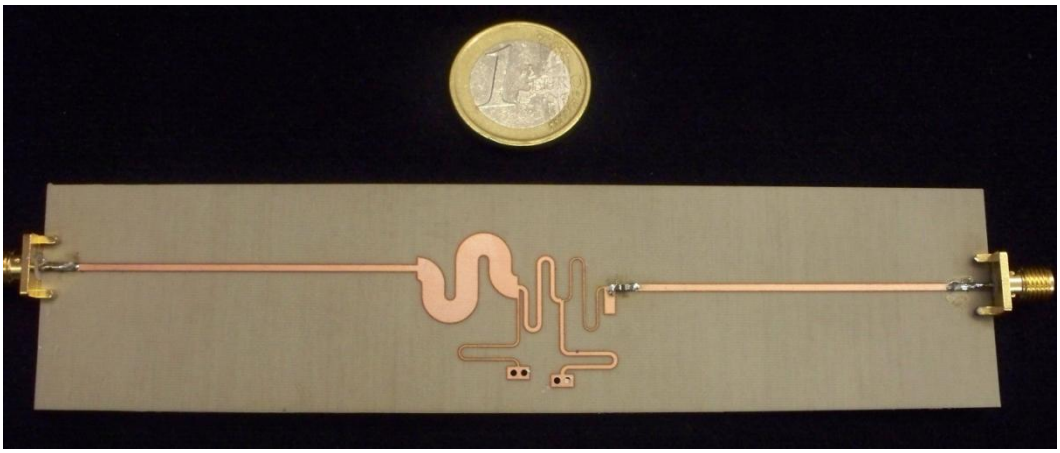


Fig. 4.7.9: matching network

If we compare the measurements results with those of the simulations, we can see that there is a considerable loss of efficiency (especially at the 2.45 GHz band). In order to try to explain this loss, we decided to fabricate and test the matching circuit and check if there is any considerable loss.

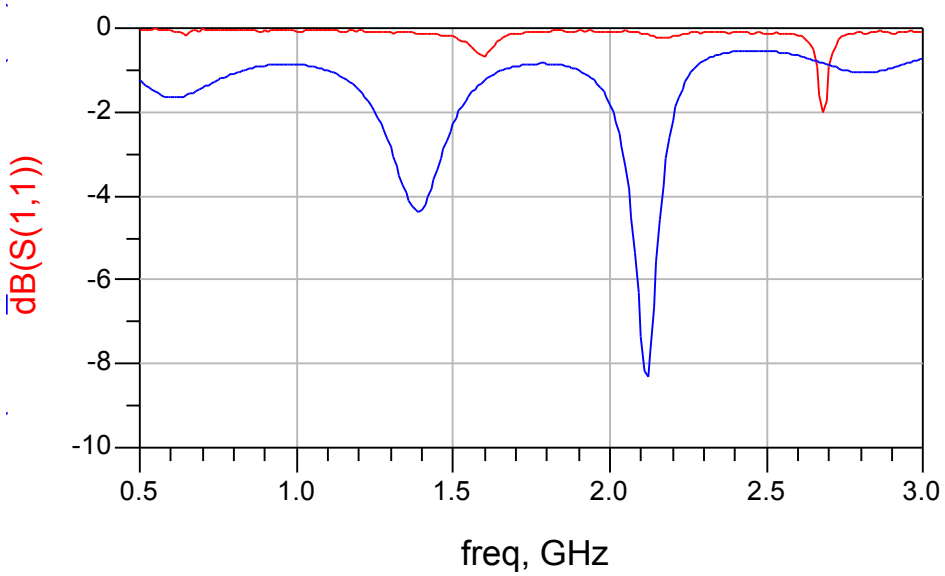


Fig. 4.7.10: S11 comparison: red line = measurement, blue line = simulation

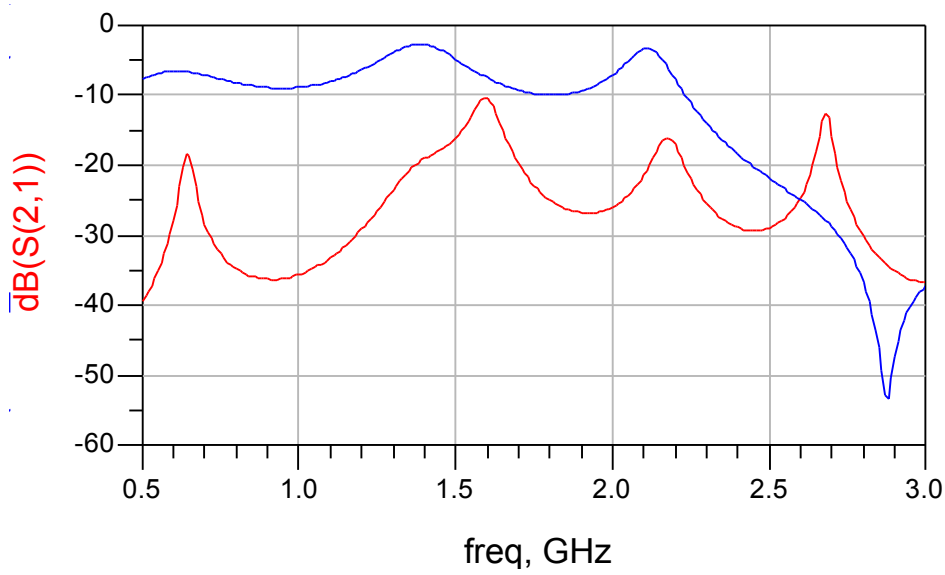


Fig. 4.7.11: S21 comparison: red line = measurement, blue line = simulation

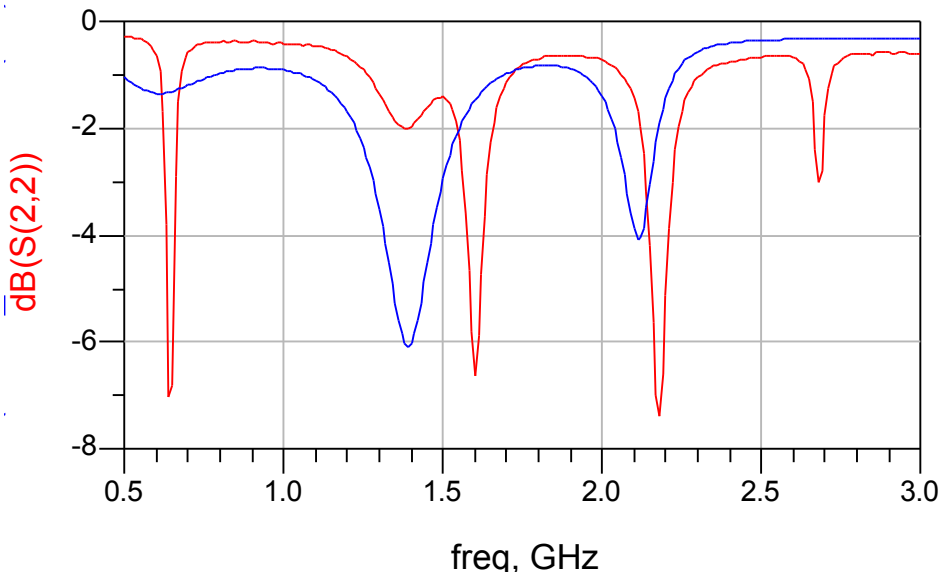


Fig. 4.7.12: S22 comparison: red line = measurement, blue line = simulation

To carry out this measurements a Rohde & Schwarz ZVA24 Vector Network Analyzer (VNA) that, after a SOLT (Short / Open / Load / Thru) calibration for a 500MHz – 3 GHz swept, was used to measure the S-parameters of the circuit. Fig. 4.7.10 – 4.7.11 – 4.7.12, show the results of the measurements.

As we expected, from Figs 4.7.10 and Fig. 4.7.11we can conclude that there is a notable difference that denotes gain loss and high refection coefficients. The reason of this behavior will be an issue of future work that will not be included in this project.

Chapter 5

Future Work and Conclusions

5.1 Future Work

Future work in the energy harvesting field is based on applying these concepts to everyday environments of a person or critical situations, where wireless powering is the only, safe, solution. In this way, it will be feasible to continuously storage energy and perform actions that, until now, conventional methods don not allow making them possible. Some possible new technologies that can be applied to this concept will be briefly described.

5.1.1 Textile Material for Antennas

Textile UWB antennas for WBAN applications have small physical size and can be easily integrated into clothing [22]. Results confirmed the good transient performance over UWB frequencies. As a conductor, high conductive metalized Nylon fabric is commonly used. Its metalized layers provide high conductivity and protection against corrosion, as well as flexibility. As dielectric substrate, acrylic fabric is often used.

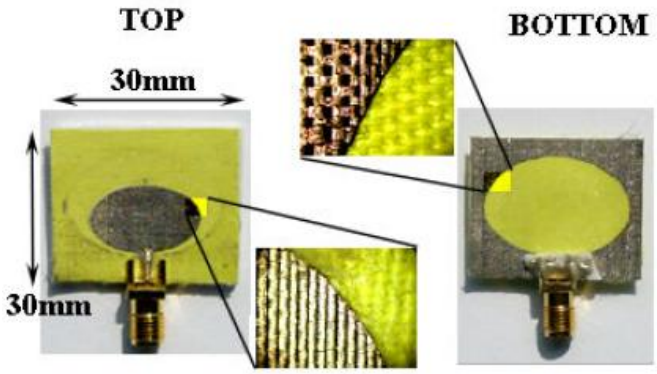


Fig. 5.1.1: textile UWB antenna [22]

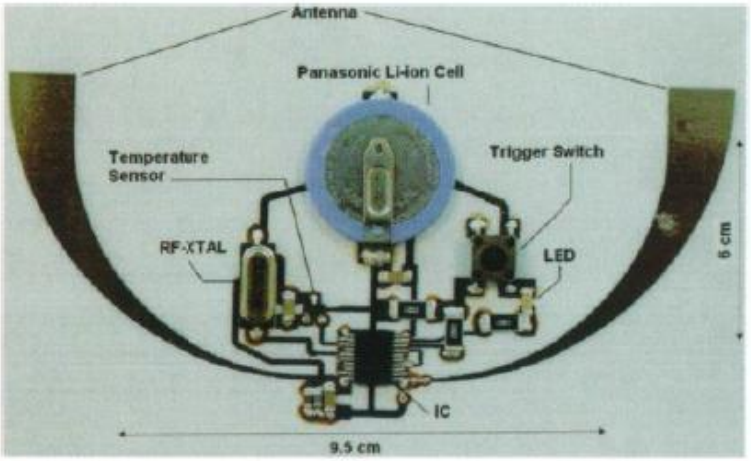


Fig. 5.1.2: wireless-sensor transmitter, on a paper substrate, using silver inkjet-printing technology [23]

5.1.2 Inkjet-Printed Antennas on Flexible Low-Cost Paper-Based Substrates

Paper, which holds one of the biggest market shares in the world, can potentially evolve the electronics market. It could eventually take the first step in creating an environmentally friendly first generation of truly "green" RF electronics and modules.

In addition, paper is one of the lowest-cost materials produced. Inkjet-printing technology (Fig. 5.1.2), which is a much faster and cleaner method than conventional wet-etching techniques that use several chemicals, can serve as a low-cost technology for fabricating RFID tags, produced in large processes on paper

5.1.3 Flexible Antenna on Polyethylene Terephthalate (PET) Substrate

These prototype antennas use low cost, wet deposition techniques to deposit silver and copper onto the substrates.

Fig. 5.1.3: Flexible folded dipole made from electro-less silver deposited on a PET substrate [24]

5.2 Conclusions

Concluded this project, we can affirm that the low power energy harvesting (in the environment) is a widespread trend that is becoming more accepted. For this new kind of "alternative energy" with the use of a rectenna, we conclude that there are still some limitations due mainly to delivered power of a mobile base station or a wireless router, is still low; this power levels can be improved using antennas arrays, or new high efficiency rectifiers. On the other hand, this energy can be stored, having as analogous application the solar energy, where sometimes photo voltaic modules are used to charge batteries that, subsequently, will supply power to other devices. We can also conclude that the use of this technology, for everyday applications, is being investigated, as well as cheaper and more flexibility material are being increasingly used at the time of the design of these applications.

Is also concluded that incident power densities, for both GSM-1900 and GSM-1800, are practically in the same order of magnitude, and that the incident power density of the ISM 2.4 GHz band, although the wireless router is closer than a mobile base station, is one order of magnitude below. We also conclude that for both, GSM and WLAN networks, the incident power density depends on the traffic at the time of measurement.

Respect to the antenna, it is concluded that the dimensions can be reduced by analyzing the current distribution on the surface of the conductors, moreover, that non-symmetrical cuts, can modify the radiation pattern of the antenna.

With respect to the rectifier, it is concluded that the use of microstrip lines for the matching circuit design, contributed to a lower soldering loss than if components, such as inductors and capacitors, were used.

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RADIO OPERATORS HANDBOOK

U.S. Marine Corps

PCN 144 000067 00

Radio Operator's Handbook

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Chapter 1

Radio Principles

Communications and information systems (CIS) are any systems whose primary functions are to collect, process, or exchange information. The fundamental requirement is to provide the Marine air-ground task force (MAGTF) commander with a reliable, secure, fast, and flexible communications network.

Communications and information systems automate routine functions, thereby freeing commanders and staffs to focus on those aspects of command and control that require experience, judgment, and intuition.

These systems and the personnel who install, operate, and maintain them play a key role in the command and control of the MAGTF. Communications and informations systems support the commander and every staff section in every phase of operations planning and execution.

These systems facilitate information flow throughout the MAGTF and provide shared situational awareness, informed decisionmaking, and rapid dissemination of decisions.

The success of the MAGTF in the modern battlespace depends on the effective employment of communications and information systems. Single-channel radio (SCR) is one of the most important components of MAGTF CIS.

Section I. Theory and Propagation

BASIC COMPONENTS OF RADIO EQUIPMENT

The radio equipment for communication between two stations and the path the signal follows through the air is called a radio link. A radio link consists of seven components: the transmitter, power supply, transmission lines, transmitting antenna, propagation path, receiving antenna, and receiver (see figure 1-1).

The transmitter generates a radio signal. The power supply (i.e., battery or generator) supplies power for the operating voltage of the radio. The transmission line delivers the signal from the transmitter to the antenna.

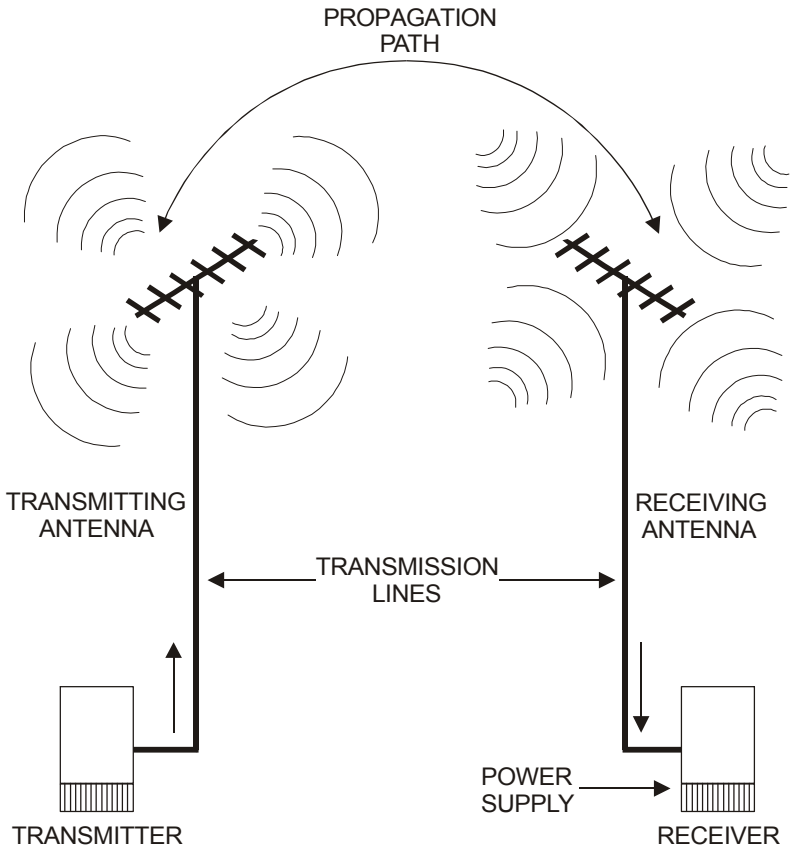

Figure 1-1. Radio Link.

The transmitting antenna sends the radio signal into space toward the receiving antenna. The path in space that the radio signal follows as it goes to the receiving antenna is the propagation path. The receiving antenna intercepts or receives the signal and sends it through a transmission line to the receiver. The receiver processes the radio signal so the human ear can hear it.

When transmitting, the radio operator aims to provide the strongest possible signal at the site of the receiving station. The best possible signal is that signal which will provide the greatest signal-to-noise ratio at the receiving antenna.

To transmit the best possible signal, select or determine the—

- Optimum frequency.
- Best antenna for that frequency based on the available space of the transmitting site.
- Proper propagation path.

RADIO WAVES

Propagation Velocity (Speed)

Radio waves travel near the surface of the Earth and radiate skyward at various angles to the Earth's surface. These electromagnetic waves travel through space at the speed of light, approximately 300,000 kilometers (km) or 186,000 miles (mi) per second.

Wavelength

Wavelength is the distance between the crest of one wave and the crest of the next wave (see figure 1-2 on page 1-4). It can also be the length of one complete cycle of the waveform. It is also the distance traveled during one complete cycle. The length of the wave is always measured in meters.

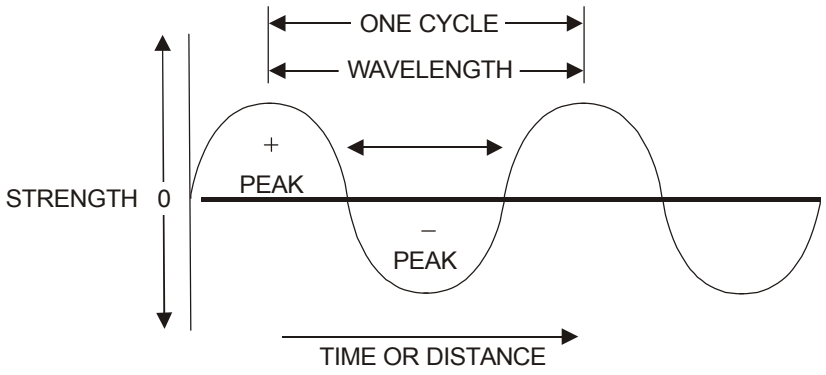

Figure 1-2. Radio Waves.

Radio Frequency

The frequency of a radio wave is the number of complete cycles that occur in one second. The longer the cycle, the longer the wavelength and the lower the frequency. The shorter the cycle, the shorter the wavelength and the higher the frequency. Frequency is measured and stated in units called hertz (Hz). One cycle per second is stated as 1 hertz. Because the frequency of a radio wave is very high, it is generally measured and stated in thousands of hertz (kilohertz [KHz]) or in millions of hertz (megahertz [MHz]). One KHz is equal to 1,000 cycles per second, and 1 MHz is equal to a million cycles per second. Sometimes frequencies are expressed in billions of hertz (gigahertz [GHz]). One GHz is equal to a billion cycles per second.

For practical purposes, the velocity of a radio wave is considered to be constant, regardless of the frequency or the amplitude of the transmitted wave. Therefore, to find the frequency when the wavelength is known, divide the velocity by the wavelength.

$$\text{Frequency (hertz)} = \frac{\text{300 million (meters per second)}}{\text{Wavelength (meters)}}$$

To find the wavelength when the frequency is known, divide the velocity by the frequency.

$$\text{Wavelength (meters)} = \frac{\text{300 million (meters per second)}}{\text{Frequency (hertz)}}$$

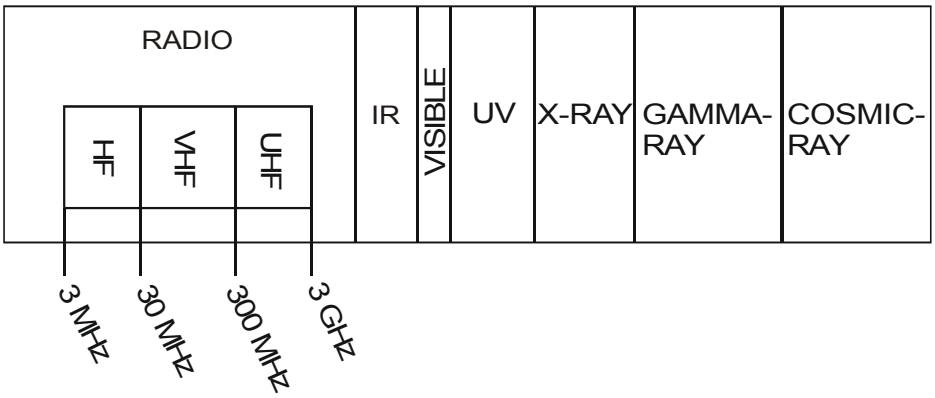

Figure 1-3. Electromagnetic Spectrum.

Within the radio frequency spectrum (see figure 1-3), radio frequencies are divided into groups or bands of frequencies. The radio frequency spectrum is part of the electromagnetic spectrum. Most tactical radio sets operate within a 2- to 400-MHz range within the frequency spectrum.

Each frequency band has certain characteristics. The ranges and power requirements shown in table 1-1 are for normal operating conditions (proper siting and antenna orientation and correct operating procedures). The ranges will change according to the condition of the propagation medium and the transmitter output power.

Table 1-1. Frequency Range Characteristics.

Band	Ground Wave Range	Sky Wave Range	Power Required
HF	0-50 miles	100-8000 miles	.5-5 kW
VHF	0-30 miles	50-150 miles	.5 or less kW
UHF	0-50 miles	N/A	.5 or less kW

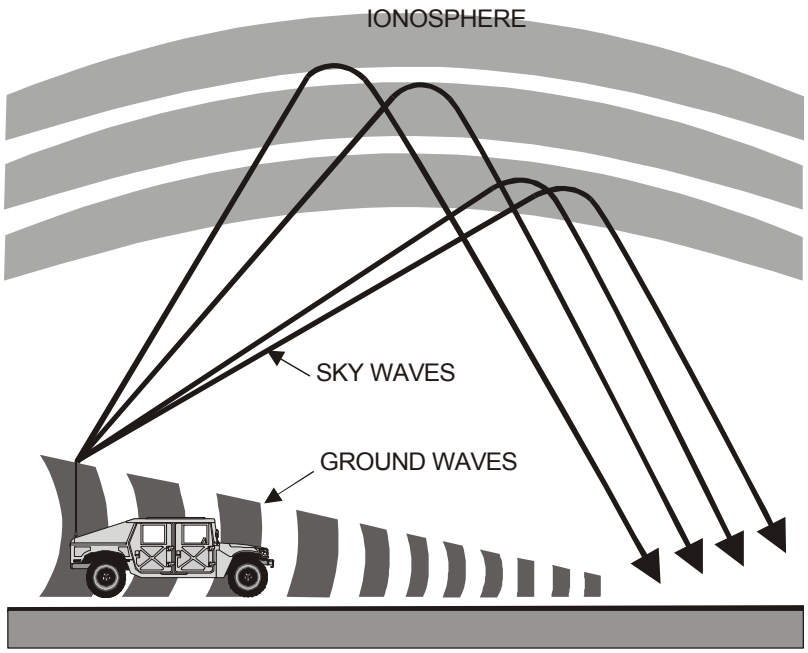

Figure 1-4. Principle Paths of Radio Waves.

RADIO WAVE PROPAGATION

There are two principal paths by which radio waves travel from a transmitter to the receiver (See figure 1-4): ground wave—which travels directly from the transmitter to the receiver and sky wave—which travels up to the ionosphere and is refracted (i.e., bent downward) back to the Earth. Short-distance, ultrahigh frequency (UHF), and upper very high frequency (VHF) transmissions are by ground waves. Long-distance, high frequency (HF) transmission is principally by sky waves. Single-channel radio sets can use ground wave or sky wave propagation for communications.

Ground Wave Propagation

Radio communications which use ground wave propagation do not use or depend on waves that are refracted from the ionosphere (i.e., sky waves). Ground wave propagation is affected by the electrical characteristics of the Earth and by the amount of diffraction (i.e., bending) of the

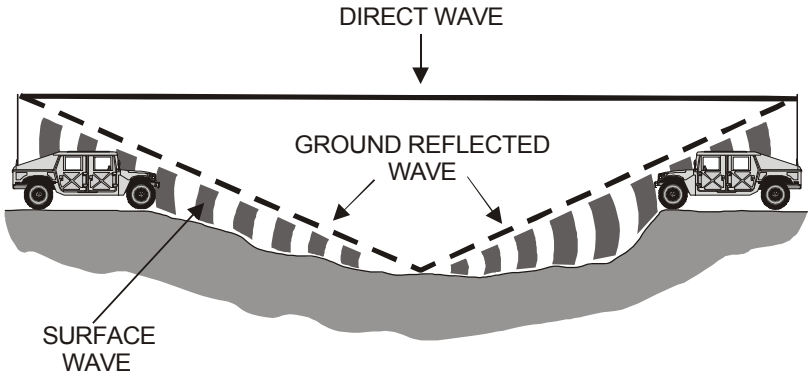

Figure 1-5. Ground Wave Propagation.

waves along the curvature of the Earth. The strength of the ground wave at the receiver depends on the power output and frequency of the transmitter, the shape and conductivity of Earth along the transmission path, and the local weather. The following paragraphs describe the components of a ground wave. See figure 1-5.

Direct Wave. The direct wave is that part of the radio wave which travels directly from the transmitting antenna to the receiving antenna. This part of the wave is limited to the line of sight (LOS) distance between the transmitting and receiving antennas, plus the small distance added by atmospheric refraction and diffraction of the wave around the curvature of the Earth. This distance can be extended by increasing the height of the transmitting antenna, the receiving antenna, or both.

Ground Reflected Wave. The ground reflected wave is that portion of the radio wave which reaches the receiving antenna after being reflected from the surface of the earth. Cancellation of the radio signal can occur when the ground reflected component and the direct wave component arrive at the receiving antenna at the same time and are 180° out of phase with each other.

Surface Wave. The surface wave, which follows the curvature of the Earth, is that part of the ground wave which is affected by the conductivity and dielectric constant of the Earth.

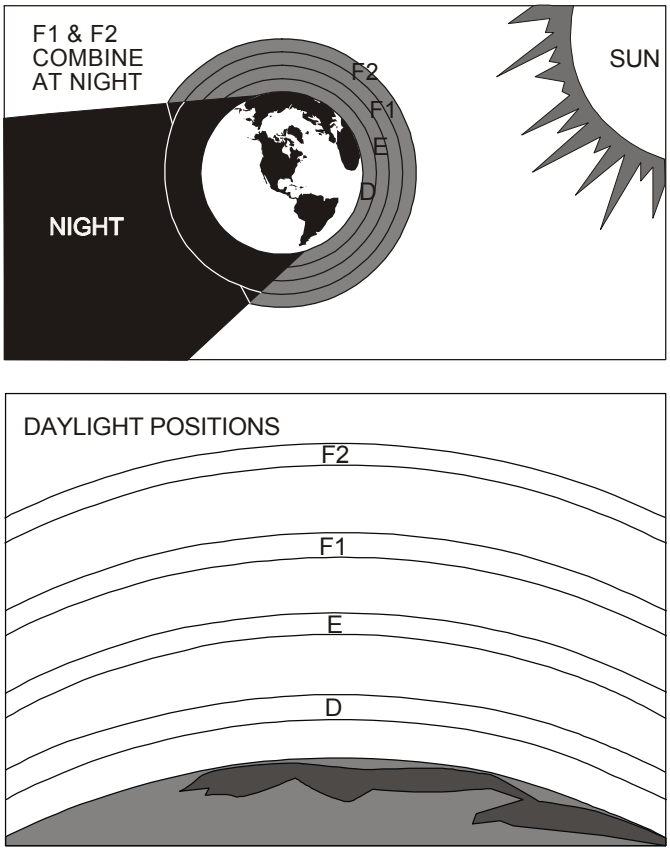

Figure 1-6. Layers of the Ionosphere.

Sky Wave Propagation

Radio communications that use sky wave propagation depend on the ionosphere to provide the signal path between the transmitting and receiving antennas.

Ionospheric Structure. The ionosphere has four layers (see fig. 1-6). In order of increasing heights and decreasing molecular densities, these layers are labeled D, E, F1, and F2. During the day, when the rays of the

Sun are directed toward that portion of the atmosphere, all four layers may be present. At night, the F1 and F2 layers seem to merge into a single F layer, and the D and E layers fade out. The actual number of layers, their height above the Earth, and their relative intensity of ionization varies constantly. The following are layers of the ionosphere:

- D—exists only during daylight hours and has little effect in bending the paths of high frequency radio waves. The main effect of the D layer is to attenuate high frequency waves when the transmission path is in sunlit regions.
- E—used during the day for high frequency radio transmission over intermediate distances (less than 2,400 km [1,500 mi]). At night, the intensity of the E layer decreases, and it becomes useless for radio transmission.
- F—exists at heights up to 380 kilometers (240 mi) above the Earth and is ionized all the time. It has two well-defined layers (F1 and F2) during the day, and one layer (i.e., F) during the night. At night, the F layer remains at a height of about 260 kilometers (170 mi) and is useful for long-range radio communications (over 2,400 km [1,500 mi]). The F2 layer is the most useful of all layers for long-range radio communications, even though its degree of ionization varies appreciably from day to day.

Factors Affecting the Ionosphere. The movements of the Earth around the Sun and changes in the Sun's activity contribute to ionospheric variations. There are two main classes of variations: regular, which is predictable; and irregular, which occurs from abnormal behavior of the Sun.

Regular Variations of the Ionosphere. The regular variations are—

- Daily—caused by the rotation of the Earth.
- Seasonal—caused by the north and south progression of the Sun.
- 27-day—caused by the rotation of the Sun on its axis.
- 11-year—caused by the sunspot activity cycle going from maximum to minimum, back to maximum levels of intensity.

Irregular Variations of the Ionosphere. The current status of the four regular variations must be anticipated when planning a communications system. There are also unpredictable, irregular variations that must be considered. They have a degrading effect (at times blocking communications) which cannot be controlled or compensated for at present. Some irregular variations are—

- Sporadic E. When it is excessively ionized, the E layer often blocks out the reflections back from the higher layers. It can also cause unexpected propagation of signals hundreds of miles beyond the normal range. This effect can occur at any time.
- Sudden ionospheric disturbance (SID). A sudden ionospheric disturbance coincides with a bright solar eruption and causes abnormal ionization of the D layer. This effect causes total absorption of all frequencies above approximately 1 MHz. It can occur without warning during daylight hours and last from a few minutes to several hours. When SID occurs, receivers seem to go dead.
- Ionospheric storms. During these storms, sky wave reception above approximately 1.5 MHz shows low intensity and is subject to a type of rapid blasting and fading called "flutter fading." These storms may last from several hours to days and usually extend over the entire Earth.

Sunspots. Sunspots generate bursts of radiation that cause high levels of ionization. The more sunspots, the greater the ionization. During periods of low sunspot activity, frequencies above 20 MHz tend to be unusable because the E and F layers are too weakly ionized to reflect signals back to earth. At the peak of the sunspot cycle, however, it is not unusual to have worldwide propagation on frequencies above 30 MHz.

Frequency Characteristics in the Ionosphere. The range of long-distance radio transmission is determined primarily by the ionization density of each layer. The higher the frequency, the greater the ionization density required to reflect radio waves back to Earth. The upper (i.e., E and F) layers reflect the higher frequencies because they are the most highly ionized. The D layer, which is the least ionized, does not reflect frequencies above approximately 500 KHz. Thus, at any given time and for each

ionized layer, there is an upper frequency limit at which radio waves sent vertically upward are reflected back to Earth. This limit is called the critical frequency.

Radio waves directed vertically at frequencies higher than the critical frequency pass through the ionized layer out into space. All radio waves directed vertically into the ionosphere at frequencies lower than the critical frequency are reflected back to Earth. Radio waves used in communications are generally directed towards the ionosphere at some oblique angle, called the angle of incidence. Radio waves at frequencies above the critical frequency will be reflected back to Earth if transmitted at angles of incidence smaller than a certain angle, called the critical angle. At the critical angle and all angles larger than the critical angle the radio waves will pass through the ionosphere if the frequency is higher than the critical frequency. When the angle of transmission becomes smaller, the radio waves will be reflected back to Earth.

Transmission Paths. The distance from the transmitting antenna to the place where the sky waves first return to Earth is called the skip distance. The skip distance is dependent on the angle of incidence, the operating frequency, and the height and density of the ionosphere. The antenna height, in relation to the operating frequency, affects the angle that transmitted radio waves strike and penetrate the ionosphere and then return to Earth. This angle of incidence can be controlled to obtain the desired area of coverage. Lowering the antenna will increase the angle of transmission and provide broad and even signal patterns in a large area. The use of near-vertical transmission paths is known as near-vertical incidence sky wave (NVIS). Raising the antenna will lower the angle of incidence. Lowering the angle of incidence can produce a skip zone in which no usable signal can be received. This area is bounded by the outer edge of usable ground wave propagation and the point nearest the antenna at which the sky wave returns to Earth. In most communications situations, the skip zone is not a desirable condition. However, low angles of incidence make long-distance communications possible.

When a transmitted wave is reflected back to the surface of the Earth, part of its energy is absorbed by the Earth. The remainder of its energy is reflected back into the ionosphere to be reflected back again. This means

of transmission—by alternately reflecting the radio wave between the ionosphere and the Earth—is called hops, and it enables radio waves to be received at great distances from the point of origin.

Maximum Usable and Lowest Usable Frequencies. There is a maximum frequency at which a radio wave will return to Earth at a given distance when a given ionized layer and a transmitting antenna with a fixed angle of radiation is used. This frequency is called the maximum usable frequency (MUF). It is the monthly median of the daily highest frequency that is predicted for sky wave transmission over a particular path at a particular hour of the day. The MUF is always higher than the critical frequency because the angle of incidence is less than 90°. If the distance between the transmitter and the receiver is increased, the maximum usable frequency will also increase. Radio waves lose some of their energy through absorption by the D layer and the portion of the E layer of the ionosphere at certain transmission frequencies.

The total absorption is less and communications are more satisfactory as higher frequencies are used up to the level of the MUF. The absorption rate is greatest for frequencies ranging from approximately 500 KHz to 2 MHz during the day. At night, the absorption rate decreases for all frequencies. As the frequency of transmission over any sky wave path is decreased from high to low frequencies, a frequency will be reached at which the received signal just overrides the level of atmospheric and other radio noise interference. This is called the lowest useful frequency (LUF) because frequencies lower than the LUF are too weak for useful communications. The LUF also depends on the power output of the transmitter as well as the transmission distance. When the LUF is greater than the MUF, no sky wave transmission is possible.

Section II. Modulation and Single Side Band Transmission

Radio communications equipment is used primarily to transmit voice and data. Although sound can be converted to audio frequency electrical energy, it is not practical to transmit it in this energy form through the Earth's atmosphere by electromagnetic radiation. For example, efficient transmission of a 20-hertz audio signal would require an antenna almost 8,000 kilometers (5,000 mi) long. This would not apply when radio frequency electrical energy is used to carry the intelligence. When radio frequency electrical energy is used, great distances can be covered; efficient antennas for radio frequencies are of practical lengths; and antenna power losses are at reasonable levels.

The frequency of the radio wave affects its propagation characteristics. In the low frequency band (.03 to.3 MHz), the ground wave is very useful for communications over great distances. The ground wave signals are quite stable and show little seasonal variation. In the medium frequency band (.3 to 3.0 MHz), the range of the ground wave varies from about 24 kilometers (15 mi) at 3 MHz, to about 640 kilometers (400 mi) at the lowest frequencies of this band. Sky wave reception is possible during the day or night at any of the lower frequencies in this band. At night, the sky wave is receivable at distances up to 12,870 kilometers (8,000 mi). In the high frequency band (3 to 30 MHz), the range of the ground wave decreases as frequency increases, and the sky waves are greatly influenced by ionospheric considerations.

In the very high frequency band (30 to 300 MHz), there is no usable ground wave and only slight refraction of sky waves by the ionosphere at the lower frequencies. The direct wave provides communications if the transmitting and receiving antennas are elevated high enough above the surface of the Earth.

In the ultrahigh frequency band (300 to 3,000 MHz), the direct wave must be used for all transmissions. Communications are limited to a short distance beyond the horizon. Lack of static and fading in these bands makes line of sight reception very satisfactory. Antennas that are highly directional can be used to concentrate the beam of radio frequency (RF) energy, thus, increasing the signal intensity.

MODULATION

Both amplitude modulation (AM) and frequency modulation (FM) transmitters produce RF carriers. The carrier is a wave of constant amplitude, frequency, and phase which can be modulated by changing its amplitude, frequency, or phase. Thus, the RF carrier "carries" intelligence by being modulated. Modulation is the process of superimposing intelligence (voice or coded signals) on the carrier.

Amplitude Modulation

Amplitude modulation is the variation of the RF power output of a transmitter at an audio rate. In other words, the RF energy increases and decreases in power according to the audio frequencies superimposed on the carrier signal.

When audio frequency signals are superimposed on the radio frequency carrier signal, additional RF signals are generated. These additional frequencies are equal to the sum of, and the difference between the audio frequencies and the radio frequency used. For example, assume a 500-KHz carrier is modulated by a 1-KHz audio tone. Two new frequencies are developed, one at 501 KHz (the sum of 500 KHz and 1 KHz) and the other at 499 KHz (the difference between 500 KHz and 1 KHz). If a complex audio signal is used instead of a single tone, two new frequencies will be set up for each of the audio frequencies involved. The new frequencies resulting from superimposing an audio frequency (AF) signal on an RF signal are called side bands.

When the RF carrier is modulated by complex tones such as speech, each separate frequency component of the modulating signal produces its own upper and lower side band frequencies. The side band that contains the sum of the RF and AF signals is called the upper side band. The side band that contains the difference between the RF and AF signals is called the lower side band.

The space occupied by a carrier and its associated side bands in the radio frequency spectrum is called a channel. In amplitude modulation, the width of the channel (bandwidth) is equal to twice the highest

modulating frequency. For example, if a 5,000 KHz (5 MHz) carrier is modulated by a band of frequencies ranging from 200 to 5,000 cycles (.2 to 5 KHz), the upper side band extends from 5,000.2 to 5,005 KHz. The lower side band extends from 4,999.8 KHz to 4,995 KHz. Thus, the bandwidth is the difference between 5,005 KHz and 4,995 KHz, a total of 10 KHz.

Frequency Modulation

Frequency modulation is the process of varying the frequency (rather than the amplitude) of the carrier signal in accordance with the variations of the modulating signals. The amplitude or power of the FM carrier does not vary during modulation.

The frequency of the carrier signal when it is not modulated is called the center or rest frequency. When a modulating signal is applied to the carrier, the carrier signal will move up and down in frequency, away from the center or rest frequency.

The amplitude of the modulating signal determines how far away from the center frequency the carrier will move. This movement of the carrier is called deviation; how far the carrier moves is called the amount of deviation. During reception of the FM signal, the amount of deviation determines the loudness or volume of the signal.

The FM signal leaving the transmitting antenna is constant in amplitude but varies in frequency according to the audio signal. As the signal travels to the receiving antenna, it picks up natural and man-made electrical noises that cause amplitude variations in the signal. All of these undesirable amplitude variations are amplified as the signal passes through successive stages of the receiver until the signal reaches a part of the receiver called the limiter. The limiter is unique to FM receivers as is the discriminator.

The limiter eliminates the amplitude variations in the signal, then passes it on to the discriminator which is sensitive to variations in the frequency of the RF wave. The resultant constant amplitude, frequency-modulated signal is then processed by the discriminator circuit which changes the

frequency variations into corresponding voltage amplitude variations. These voltage variations reproduce the original modulating signal in a headset, loudspeaker, or teletypewriter. Frequency modulation is generally used by radiotelephone transmitters operating in the VHF and higher frequency bands.

SINGLE SIDE BAND TRANSMISSION

The intelligence of an AM signal is contained solely in the side bands. Each side band contains all the intelligence needed for communications. Therefore, one side band and the carrier signal can be eliminated. This is the principle on which single side band (SSB) communications is based. Although both side bands are generated within the modulation circuitry of the SSB radio set, the carrier and one side band are removed before any signal is transmitted.

The side band that is higher in frequency than the carrier is called the upper side band (USB). The side band that is lower in frequency than the carrier is called the lower side band (LSB). Either side band can be used for communications as long as both the transmitter and the receiver are adjusted to the same side band. Most SSB equipment operates in the USB mode. The transmission of only one side band leaves open that portion of the RF spectrum normally occupied by the other side band of an AM signal. This allows more emitters to be used within a given frequency range.

Single side band transmission is used in applications when it is desired to—

- Obtain greater reliability.
- Limit size and weight of equipment.
- Increase effective output without increasing antenna voltage.
- Operate a large number of radio sets without heterodyne interference (e.g., whistles and squeals) from radio frequency carriers.
- Operate over long ranges without loss of intelligibility because of selective fading.

Chapter 2

Single-Channel Radio

SINGLE-CHANNEL RADIO COMMUNICATIONS EQUIPMENT

Single-channel radio is the principal means of communications support for MAGTF maneuver units. SCR communications equipment is easy to operate. The networks are easily established, rapidly reconfigured, and, most importantly, easily maintained on the move. SCR provides secure voice communications and supports limited data information exchange. SCR in the VHF and UHF bands is normally limited to line of sight. In the HF band, SCR can support long-range communications. SCR satellite communications (SATCOM) provides mobility, flexibility, and ease of operation with unlimited range. Limitations of SCR include susceptibility to enemy electronic warfare (i.e., cosite, terrain, and atmospheric interference); the requirement for close coordination and detailed planning (i.e., a need for common timing, frequency, and equipment); and limited spectrum availability. The latter is particularly critical in the case of SATCOM.

MAGTF SCR equipment is fielded in many configurations and includes hand-held, manpack, vehicle-mounted, bench-mounted, and sheltered radios. These radios operate in simplex and half-duplex modes. The most widely employed tactical radios provide integrated communications security (COMSEC) and jam resistance through frequency hopping. Tactical SCRs operate in the three military radio frequency bands shown in Table 2-1 on page 1-2.

Table 2-1. SCR Equipment.

Frequency Band	MAGTF SCR Equipment Used	Operating Frequency Range	Typical Application
HF	AN/PRC-104 AN/GRC-193 AN/MRC-138 AN/TSC-120	2-29.999 MHz	Radio line of sight and beyond/long range
VHF	AN/VRC-12 family: AN/PRC-68 AN/PRC-77 SINCGARS family: AN/PRC-119 AN/VRC-88 (A, D) AN/VRC-89 (A, D) AN/VRC-90 (A, D) AN/VRC-91 (A, D) AN/VRC-92 (A, D) AN/GRC-213 AN/MRC-145	30-88 MHz	Radio line of sight and relay/retrans-mission
	AN/PRC-113 AN/VRC-83	116-150 MHz	Critical line of sight (ground to air)
UHF	AN/PRC-113 AN/VRC-83 AN/GRC-171	225-400 MHz	Critical line of sight (ground to air)
	AN/PSC-3 AN/PSC-5		SATCOM footprint

HIGH FREQUENCY RADIO

HF radio equipment is capable of both long- and short-range secure voice and data communications. Data communications capability is typically limited to rates of 2.4 kilobits per second (kbps). Data transmission requires modems specifically designed for operation in this band of the radio spectrum. The AN/PRC-104 is capable of remote operation by using the analog AN/GRA-39B radio remote control. See fig. 2-1.

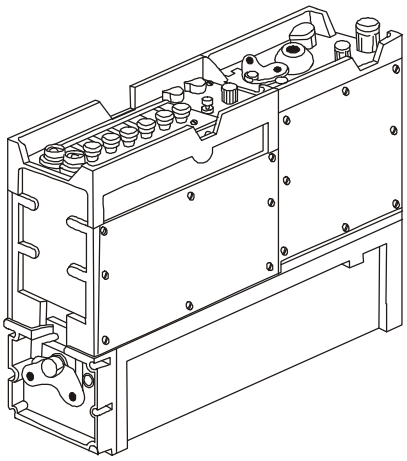

Figure 2-1. AN/PRC-104 HF Radio.

High frequency communications are capable of traveling around the world under the right conditions. This accounts for the large number of signals and noise in the receiver (e.g., thunderstorms). Conversely, the HF transmission may be intercepted and traced by the enemy who is many hundreds of miles away. VHF and UHF communications are normally limited to line of sight; therefore, their range is restricted. UHF transmissions may also be used in satellite communications, increasing ranges to thousands of miles.

High Frequency Radio Employment Considerations.

The primary advantage of using HF radio is its capability to provide long-range, over the horizon (OTH) communication. Successful data communications over the HF range depends on several factors: equipment siting, proper equipment grounding, types of antennas used, and other considerations such as tactical employment of radio equipment, path assessment and analysis, and frequency planning and assignment. When commercial data terminal equipment (DTE) is used, users employing HF radio equipment need to be aware of radio interference and potential shock hazards that can easily affect unprotected DTE. Whenever possible, HF radio equipment should be remoted from DTE.

High Frequency Radio Environmental Limitations

The primary limiting factors when using HF radios are frequency allocation and management and bandwidth availability. Frequency allocation and management is concerned with frequency, time of day, time of year, and location. The ability to reflect HF radio waves off the ionosphere to a distant location is in a constant state of flux because of activity in the ionosphere. The Sun's radiation causes disturbances in the ionosphere, with most changes taking place in what is known as the F layer (see chapter 1 for more details). Sunrise and sunset can be the most difficult times for HF communications. The F layer splits into two separate layers around sunrise and recombines into one layer around sunset. These splits affect transmission distances as the area "skipped over" increases and decreases. At times, solar storms can eliminate all HF communications. HF transmission paths must be constantly monitored to achieve a dependable HF link. HF radio data communications capabilities are limited by the bandwidth that is imposed by legal constraints and the physics of the spectrum. The bandwidth available in the HF spectrum limits the channel bandwidth, which limits data throughput.

High Frequency Propagation

There are two modes of propagation in HF: ground wave and sky wave. See figure 2-2.

Ground Wave. Ground wave propagation involves the transmission of a signal along the surface of the ground. The maximum ground wave range for most tactical HF communications is about 20 to 30 kilometers (12 to 22 miles) for manpack equipment and 80 to 100 kilometers for high-power vehicular and van equipment. The range may be decreased by heavy vegetation (e.g., Camp Lejeune), mountainous terrain (e.g., Camp Pendleton), or dry desert soil (e.g., Twenty-nine Palms). A ground wave circuit will generally be free of fading and may last for the entire 24-hour period without the need to change frequencies.

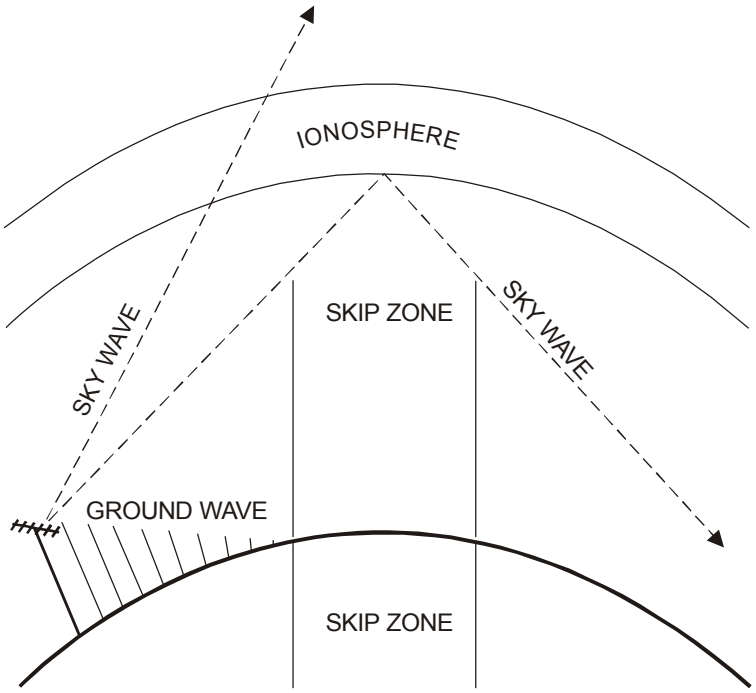

Figure 2-2. HF Propagation.

Sky Wave. Beyond this range, it is necessary to communicate by sky wave. Sky wave propagation involves the bending of the signal by the ionosphere. Frequencies are very important, as those above a certain value will not bend back to earth but will punch through the ionosphere into outer space. On the other hand, lower frequencies are noisier and become absorbed by the ionosphere. The reflective nature of the ionosphere will change when sunlight hits it each day. As a result, at least two frequencies are usually required during a 24-hour period: a low, night frequency and a higher, day frequency.

Skip Zone

A skip zone is where no signals will be received from a particular transmitter for a particular frequency. Skip zones are formed when the nearest point at which a sky wave is received is beyond the furthest point at which a ground wave is received. By using an antenna with a high

radiation take-off angle (i.e., the angle measured from the Earth's surface to horizon up to the direction of propagation towards the ionosphere), HF radio waves can be bounced off the ionosphere and come back to earth closer than they can with more commonly used antennas. This can cause the skip zone to disappear if the waves do not punch through.

The use of high radiation take-off angles is called near-vertical incident sky wave (NVIS) communications. The limit of the effective range of NVIS communications is usually about 300 miles. NVIS communications require a horizontally polarized antenna and are done over frequencies between 2 and 12 MHz. Launch angles can be changed by altering the antenna's height above ground, but for most tactical applications one-quarter wavelength above ground is sufficient. NVIS communications are particularly useful because they can be transmitted from moving vehicles. Used correctly, NVIS provides reliable, continuous communications beyond the range of HF ground wave and VHF and UHF line of sight. Multipath interference occurs when both the sky wave and the ground wave signals from the transmitter arrive at different times at the receiver. See figure 2-3. More detailed information on HF propagation and antennas may be found in MCRP 6-22D, *Antenna Handbook*.

VERY HIGH FREQUENCY RADIO

The primary MAGTF VHF radio is the single-channel ground and airborne radio system (SINCGARS). SINCGARS is a family of lightweight combat radios that serves as the primary means of communications for command and control and fire support on the battlefield. SINCGARS is the standard VHF-FM tactical radio for the Marine Corps, replacing the AN/PRC-77 and the AN/VRC-12 family. The system provides high security against threat electronic warfare (EW) by using frequency hopping with integrated COMSEC. It is capable of voice and data transmission (up to 16 kbps under optimum conditions and over limited distances) over the VHF-FM frequency range of 30 to 87.975 MHz. See MCRP 6-22A, *Talk II SINCGARS Multiservice Communications Procedures for the Single-Channel Ground and Airborne Radio System*, for more details.

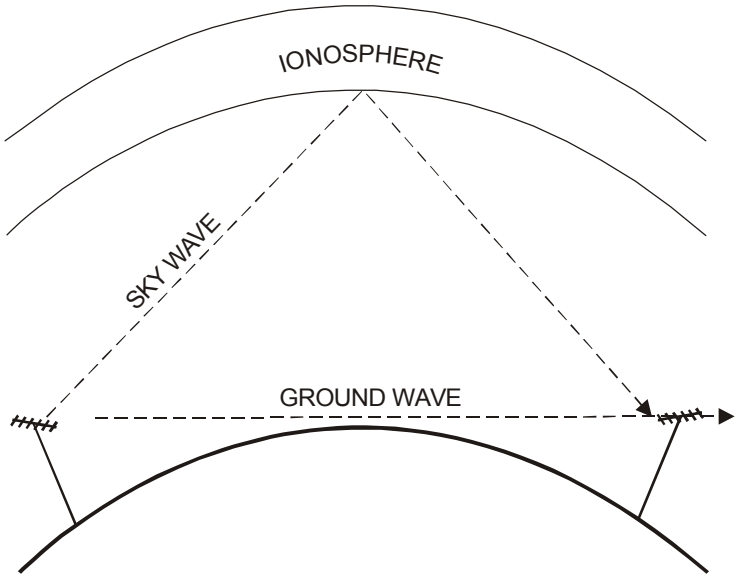

Figure 2-3. Multipath Interference.

There are seven different SINCGARS configurations available, depending on the requirements of the user. These configurations include the manpack AN/PRC-119 (see figure 2-4 on page 2-8), typically used in infantry operations, and vehicle-mounted variants. The radio provides voice communications ranges of up to 8 km for the manpack and 35 km for vehicular configurations. SINCGARS is capable of remote operation by using the analog AN/GRA-39B radio remote control, the digital HYX-57 wire-line ADAPTER, or the digital C-11561 (C)/U remote control unit (RCU).

The SINCGARS radio has undergone a systems improvement program (SIP). This radio is referred to as the SINCGARS SIP. The primary improvements relate to the data transmission capabilities of the system. A forward error correction appliqué was implemented in the receiver and/or transmitter, and a new packet data mode was created to better support packet networks. In addition, an improved channel access protocol was added, which optimizes data throughput performance while minimizing impact on voice communications on the same SINCGARS channel.

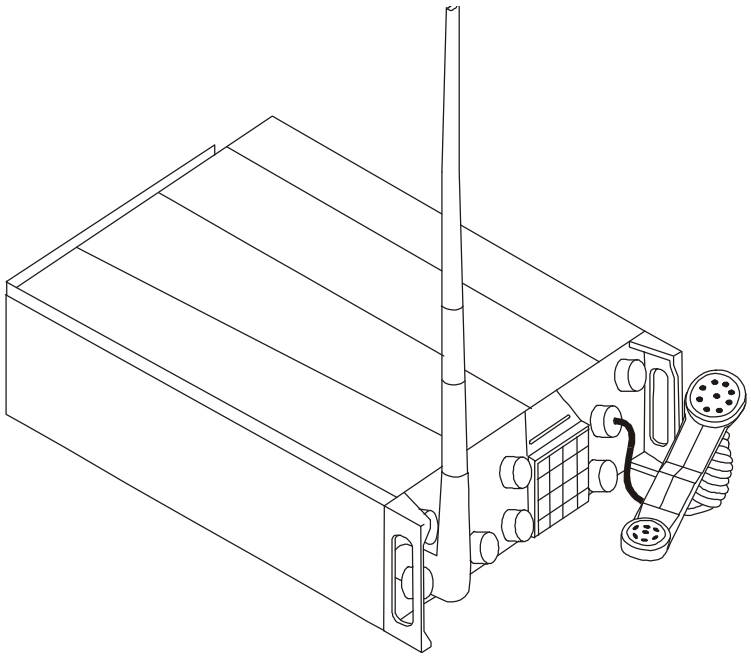

Figure 2-4. AN/PRC-119 SINCGARS Radio.

THE SINCGARS SIP radio is also available in a downsized version—the result of an advanced systems improvement program (ASIP). This radio is referred to as the SINCGARS ASIP. This radio will retain all the functionality of the full-size SIP radio but is half the size. It weighs 7.6 pounds (including the battery). The radio is interchangeable with previous SINCGARS versions, including the capability to be mounted in older vehicular adapter assemblies. A new feature of the SINCGARS ASIP provides a retransmission capability while operating in the packet data mode and will also employ a new, fast-channel access protocol for improved operations in shared voice or data nets.

The AN/ARC-210 multipurpose radio supports single-channel air-to-air, air-to-ground, and ground-to-air communications in tactical Navy and Marine Corps fixed- and rotary wing aircraft. It can transmit and receive VHF-FM, VHF-AM, and UHF signals. It is compatible with SINCGARS, HAVE QUICK, and HAVE QUICK II frequency hopping UHF radios, and it can accept 25 preset, single-channel frequencies. The AN/ARC-210 requires a TSEC/KY-58 encryption device to encrypt transmissions and decrypt received signals.

Hand-Held Very High Frequency Radios

Radio operators may have the opportunity to use various commercial off-the-shelf (COTS) VHF radios in the Fleet Marine Force (FMF). All of these radios have been open-purchased by the user units and are not part of the official Marine Corps table of equipment. Therefore, they have not been assigned a table of authorization material control number (TAMCN).

These hand-held radios are typically small, lightweight, battery-powered equipment which provides clear (and in some cases secure) voice communications on up to 100 different channels. Some models come with headsets and microphones. Hand-held radios are mostly used at the infantry-squad level or in maritime prepositioning force (MPF) offloads.

Very High Frequency Radio Employment Considerations

Operator maintenance of the radio equipment, antennas, cable assemblies, and equipment grounding as well as proper planning and selection are essential to reliable communications. Frequency separation, radio antenna separation, remote rekeying when using COMSEC, and power output are significant employment factors. SINCGARS may be limited to the single-channel mode when operating with some Navy ships. When SINCGARS is employed in the frequency hopping mode, the following operating factors need to be taken into account: hopset (i.e., frequency segment allocation), net sych time and mission date, antenna placement (cosite interference is more of a concern than in the single-channel operating mode), and power setting. SINCGARS radios configured for different hopsets that dial into the same numbered net will not be able to communicate. MCRP 6-22A provides detailed information on the employment of SINCGARS.

VHF SCR is the primary communications system for combat and combat support units while on the move. The predominant mode of operation is secure voice. However, use of VHF radio for data communications will increase with the fielding of tactical information systems at the battalion level and below. Small, hand-held VHF radios are used at the small-unit level in the MAGTF. These radios are often commercial items

that lack compatibility with SINCGARS and do not have integrated COMSEC. Their use should be governed accordingly.

Very High Frequency Radio Environmental Limitations

The primary limiting factors when using VHF radios are range and frequency availability. VHF radios can provide reliable communications for ranges of up to 10 miles, depending on the equipment operating constraints and the operating environment. Unit location must be considered when employing radios that operate in the VHF spectrum. Most circuits are limited to radio line of sight, known as four-thirds earth curvature. VHF radio signals essentially follow the curvature of the earth to a distance that is approximately one-third greater than the distance to the horizon. Foliage interferes with VHF signals and may reduce normal operating ranges to significantly less than 10 miles.

Very High Frequency Propagation

Radios in the SINCGARS family are the principal VHF transceivers used by the Marine Corps. The mode of communications used in this range is frequently referred to as frequency modulation. VHF will extend slightly beyond line of sight due to diffraction or bending of the signal by the atmosphere (see fig. 2-5). At frequencies in the 30-MHz range, VHF will often act like HF ground wave. The range of reliable communications is generally no more than 15 to 20 kilometers (9.3 to 12.4 mi) under normal field conditions for manpack equipment. Vehicle-mounted equipment may communicate farther because of higher transmitter power and better antennas.

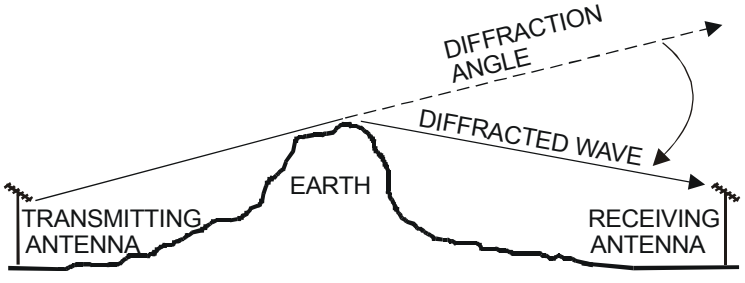

Figure 2-5. VHF Diffraction.

VHF LOS can also be plagued by multipath interference when the direct ray and a reflected ray traveling over a slightly longer path combine at the receiver antenna so that they periodically cancel or reinforce each other (see fig. 2-6). The signal fades in and out over a period of time as a result. Fading is not as great a problem with immobile equipment because corrective action can be taken, but fading can cause significant problems when one or more of the units are mobile.

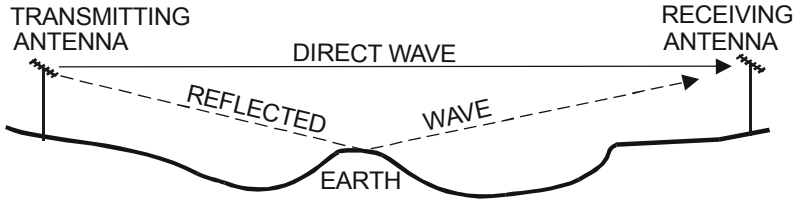

Figure 2-6. VHF LOS.

ULTRAHIGH FREQUENCY RADIO

Military UHF radio equipment operates in the 116 to 150 MHz upper-VHF frequency range and the 225 to 400 MHz military UHF radio spectrum. MAGTF UHF radio sets such as the AN/PRC-113 (see figure 2-7 on page 2-12) are capable of data communications at 16 kbps under optimal conditions. MAGTF ground and airborne UHF radios incorporate the HAVE QUICK Electronic Counter-Counter Measures capability and operate in single-channel and frequency hopping modes. The HAVE QUICK UHF radio is capable of remote operation by using the AN/GRA-39B or HYX-57.

Ultrahigh Frequency Radio Employment Considerations

UHF radios are used for forward air control (FAC) ground-to-air communication. Line of sight between radios is critical for reliable communications. Significant range differences are encountered between UHF radios employed for ground-to-air and ground-to-ground communications. Greater range is achieved when employed from ground-to-air because of the increased line of sight. When UHF radios are employed in

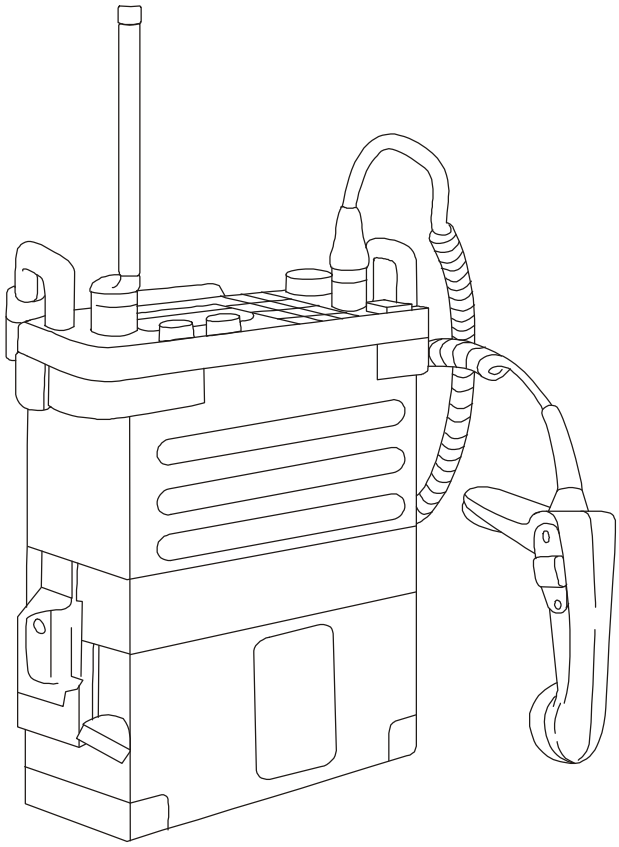

Figure 2-7. AN/PRC-113. UHF Radio.

the frequency hopping mode, the following operating factors must be understood for proper operation: hopset, time of day, antenna placement, and power setting.

Ultrahigh Frequency Radio Environmental Limitations

The primary limiting factor when using UHF radios is range (i.e., critical line of sight). Critical line of sight can be described as "what you see is what you get." As long as the radio's antenna has optical line of sight to another radio's antenna, the two will be able to transmit and receive. For this reason, UHF radios are used primarily in air-to-ground communications.

Ultrahigh Frequency-Tactical Satellite

The AN/PSC-5 is a portable, battery-operated, half-duplex UHF transceiver. It is employed for long-range communications. It weighs approximately 14 pounds including antenna and batteries. The AN/PSC-5 provides two-way voice and data communications by satellite. It operates on the UHF frequency band of 225- to 400-MHz range. It provides 2,400 to 16,000-bits per second (bps) data rate, depending on mode setting. Only one operator is required to operate it. The United States Marine Corps UHF tactical SATCOM system supports and augments the high precedence command and control and common-user, single-channel requirements of a Marine air-ground task force and its major subordinate headquarters.

The space segments used by the AN/PSC-5 are the Fleet Satellite Communications, leased satellite communications, and UHF follow-on satellites. All the satellites are located in geosynchronous orbits and permit interconnections among mobile, ground terminals. The one-way distance to servicing satellites is approximately 25,000 miles, resulting in a round-trip propagation delay of approximately one-quarter of a second. The shape of the satellite footprints is roughly circular but elongated from north to south. This is caused by the angle at which the signal hits the Earth's surface and by the curvature of the Earth's surface.

Multiple-access schemes can operate either with fixed-channel assignments to the various users or with channels being assigned in varying fashion according to demand. The latter is called demand assigned multiple access (DAMA). With demand assignment, the user makes a channel request, and a channel is allocated after a brief time lag. The DAMA scheme of operation is employed on UHF-tactical satellite (TACSAT) to share available channels more efficiently. The radio systems are compatible with the KY-57 (wideband mode only), the KY-99 and ANDVT (narrowband mode only), and the KG-84C (wideband or narrowband) COMSEC equipment. This radio equipment is also capable of remote operations by using the AN/GRA-39B (narrowband mode) or HYX-57 (wide-band mode).

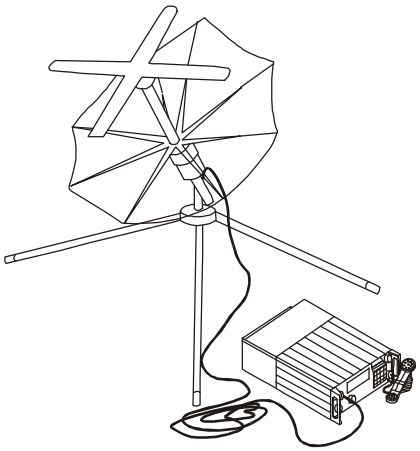

Figure 2-8. AN/PSC-5 UHF TACSAT Radio.

TACSAT Radio Employment Considerations. Because of TACSAT's limited availability, the MAGTF employs TACSAT primarily to support critical, long-range communications requirements (e.g., communications support for deep reconnaissance operations or connectivity to the tactical echelon of a MEU[SOC] when deployed ashore). The AN/PSC-5 is the primary DAMA-capable, TACSAT radio available to the MAGTF (see fig. 2-8). TACSAT limitations include the competition for available frequency resources and channel time on the satellite. If only narrow band channels are available, channel-data rates are limited to 2,400 bps. Channel congestion, noise, and network saturation will affect the information flow on satellite channels and will require a significant reduction in the data transmission rates to sustain data communications. Transmit power selection can be critical. Increasing the transmit power can decrease net effectiveness. Larger directional antennas provide increased signal gain, which increases the transmitted signal power. Antennas for these systems are lightweight and fragile and, therefore, require constant maintenance and inspection for proper operation. Satellites are shared resources. Exact frequency, bandwidth, and power of every carrier transmitted through the satellite is strictly controlled by a higher authority.

TACSAT Radio Environmental Limitations. The primary environmental limitations on TACSAT radios are signal propagation delay, location on the Earth, terrain masking, and weather effects. Timing between DTE can be a critical factor in SATCOM because the satellite, acting as a relay between radios, is about 25,000 miles away. There is

approximately a one-fourth second propagation delay between sending and receiving stations. This delay can interfere with systems that automatically retransmit if an acknowledgment is not received after a very short time-out period. As unit location changes, the "look angle" (i.e., angle above the horizon) to the satellite can affect net reliability. The orbit of a satellite allows it to cover a certain footprint on the earth. Satellites in equatorial orbit can cover large portions of the Earth, both north and south of the equator, but as the user moves closer to the Earth's poles, the TACSAT terminal may exceed the satellite footprint. This will cause intermittent or lost communications. Terrain can also have this effect by interfering with the satellite and TACSAT terminal line of sight. Thunderstorms, heavy snowstorms, and hail also affect satellite transmissions by damaging antennas and changing the electromagnetic environment.

Ultrahigh Frequency Propagation

UHF frequency propagation is used for ground-to-air, air-to-air, satellite, and tactical multi-channel communications. Communications are limited to LOS but may extend for more than 500 kilometers as long as the aircraft is high enough to be within LOS (see fig. 2-9). It is even possible for UHF communications ranges to a satellite to be more than 35,000 kilometers.

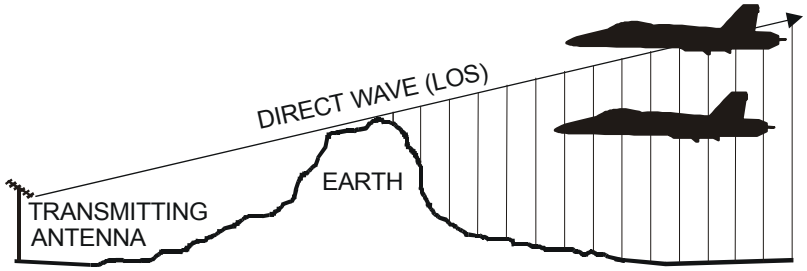

Figure 2-9. UHF LOS.

DATA COMMUNICATIONS

SCR can also transmit and receive data by using terminal devices such as the digital message system (DMS)—previously called and more commonly known as the digital communications terminal (DCT)—and the tactical communications interface module (TCIM). The DCT, data

automated communications terminal (DACT), and the TCIM are critical in enabling data communications at the tactical level over SCRs.

Digital Message System-AN/PSC-2

The DMS-AN/PSC-2 is a hand-held communications device that can be operated with either a standard military radio or telephone field wire equipment. The DMS is used to compose, edit, display, transmit, and receive information. By menu selection, formatted text, free text, and digitized map messages are transmitted over tactical communications equipment. DMS is being used directly for air support, fire support coordination, reconnaissance, medical evacuation, and other functions. DMS uses a burst transmission capability which reduces the vulnerability to enemy radio direction finding and jamming.

Data Automated Communications Terminal

The DACT is a small, tactical computer and communications terminal which gives users the capability to receive, process, and transmit various messages, to include text and symbology, used by tactical data systems. The DACT will effectively replace the DCT when it achieves full operational capability in FY 03, and it will provide much greater functionality below battalion levels. This will include an embedded global positioning system (GPS) receiver, the ability to share a common picture of the battlespace, automated data exchange, and MAGTF command, control, communications, computers, and intelligence (C4I) network connectivity. The DACT will be transportable by foot-mobile Marines and mounted in tactical or armored vehicles.

Tactical Communications Interface Module

The TCIM provides the communications link between the tactical computers of the communications and information systems within the MAGTF and the local and wide area networks, switched backbone (SBB), and radio nets. There are two versions of the TCIM card: an internal personal computer asynchronous transfer card to mount directly in the computer, and an external version with the card mounted in a portable chassis. TCIM software was developed for open-systems architectures. Small computer systems interfaces (SCSI) provide interoperability between the TCIM and other Marine Corps computers.

Chapter 3

Equipment Siting and Grounding Techniques

Two factors play an important role in equipment siting: optimum communications and camouflage. It is often difficult to find communications sites which are hidden from enemy view, fire, and direction finding and afford good communications connectivity. The ideal location for a radio antenna is as far away from cover as possible, such as a bare mountain top or in the middle of a large field. However, this goes against the commander's tactical requirement for troops and equipment to be camouflaged and concealed as much as possible. Therefore, planning the location of equipment must be detailed to achieve the best results. See Appendix A for a review on topographical maps and grid coordinates.

HIGH FREQUENCY

In the presence of hills (without large trees), the following guidelines for ground wave links should be used:

- Locate HF antennas just below the top of the hill in the direction of desired communications. Often the signal will be greater below than on the top. This will also minimize interference and/or jamming from the opposite direction.
- Move the antenna back from the hill if a hill is between the operator and the distant station with which the operator wishes to communicate. The signal strength can vary widely in the region immediately behind a hill. If it is necessary to set up behind a hill, then it may also be necessary to set up a variety of antennas located at different distances from the hilltop to see which one offers the best performance.

Long-distance, HF sky wave signals of more than several hundred kilometers are often best transmitted and received at angles just above the horizon level. Obstacles on the horizon will cause the signal to travel a

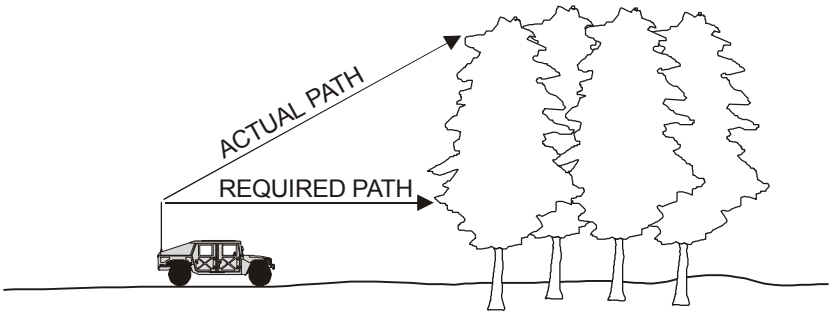

Figure 3-1. Low HF Horizon Angles.

higher path angle and may reduce the circuit reliability as a result (see fig. 3-1). Wire fences between the operator and the horizon will also lessen the chances of getting through (see fig. 3-2).

An HF ground wave signal will follow the terrain much better than higher frequency signals. It will be weakened by trees (more so when

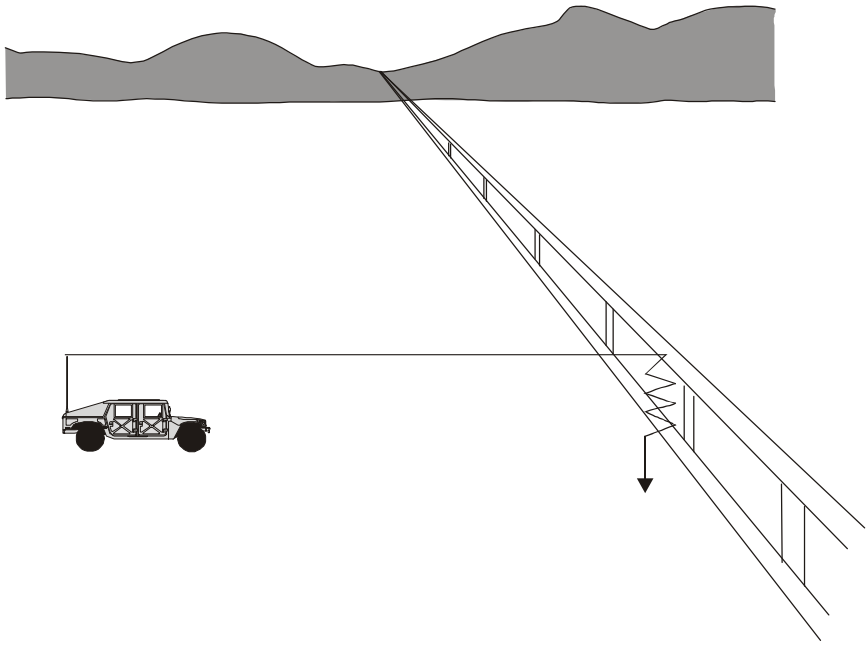

Figure 3-2. Effect of Wire Fences and Power Lines.

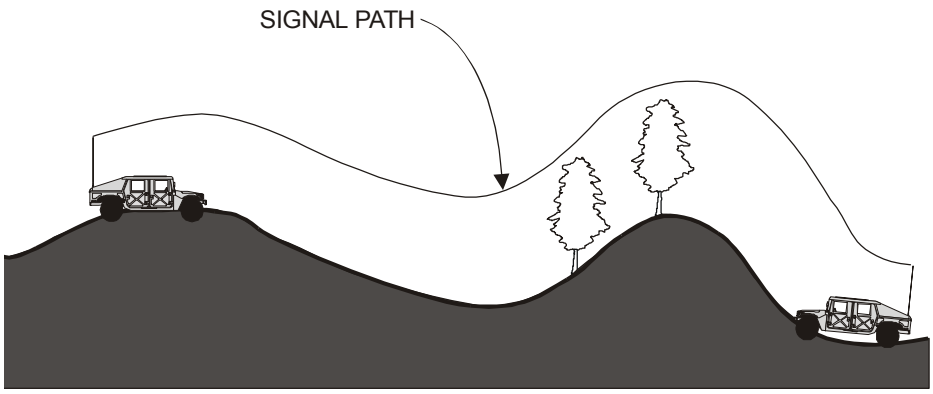

Figure 3-3. HF Ground Wave Path.

they have leaves) and rugged terrain, but the signal may still get through (see fig. 3-3).

VERY HIGH FREQUENCY AND ULTRAHIGH FREQUENCY

Obstacles such as trees, buildings, and hills between a transmitter and receiver will weaken the signal or stop it. Aircraft flying along the path will also interfere with reception (see fig. 3-4). A clear signal path

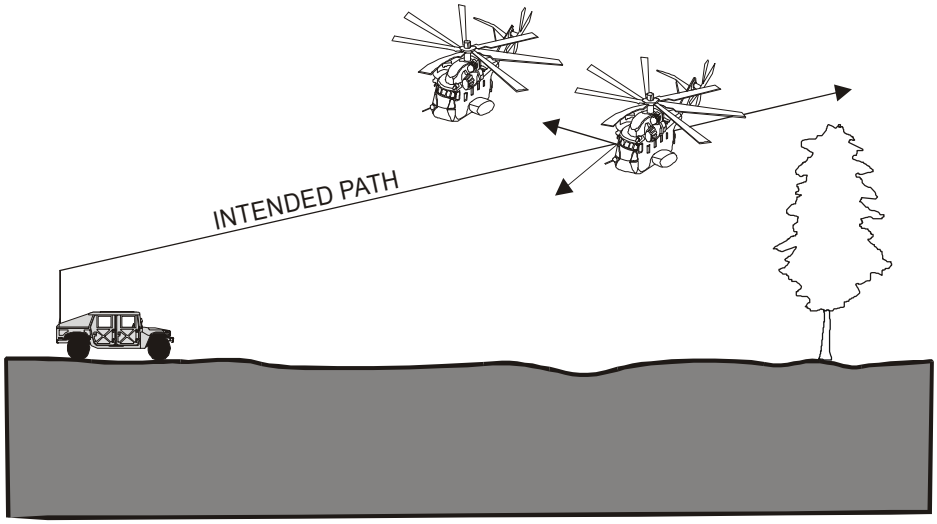

Figure 3-4. Aircraft Along Signal Path.

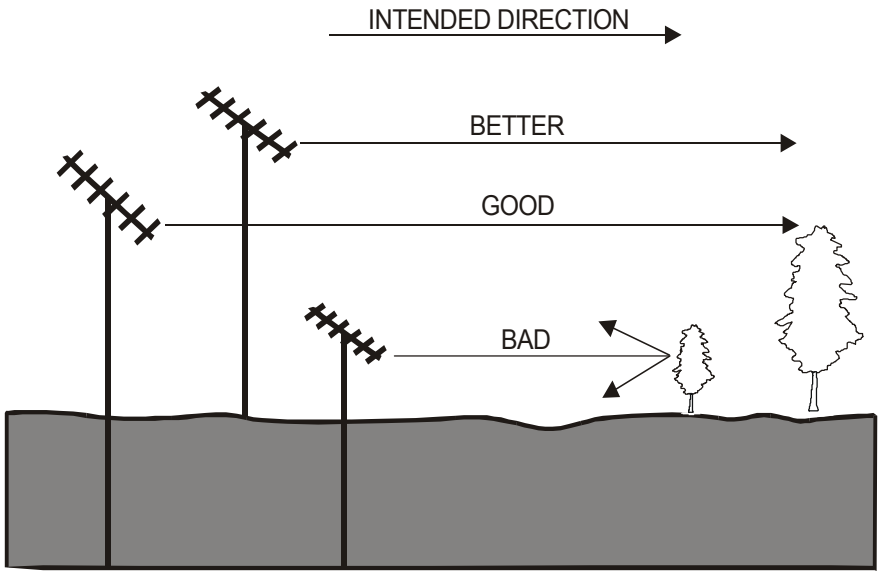

Figure 3-5. Antenna Obstacles.

between the transmitter and the intended receiver, especially for LOS communications, is preferred. The antenna must be positioned as high as possible to overcome obstacles, especially if communication is in the direction of trees or buildings (see fig. 3-5). Keep equipment as far back as possible from obstacles in the direction of the signal path to prevent interference or damage to equipment.

In some situations, solid obstacles may actually improve a link by providing a sharper surface to diffract over or reflect from (see fig. 3-6).

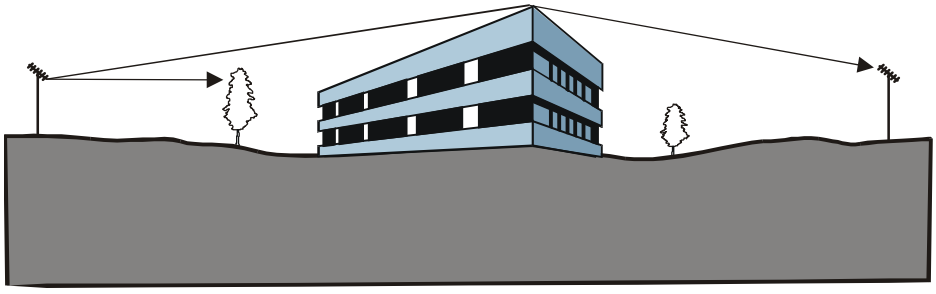

Figure 3-6. Diffraction Over Building.

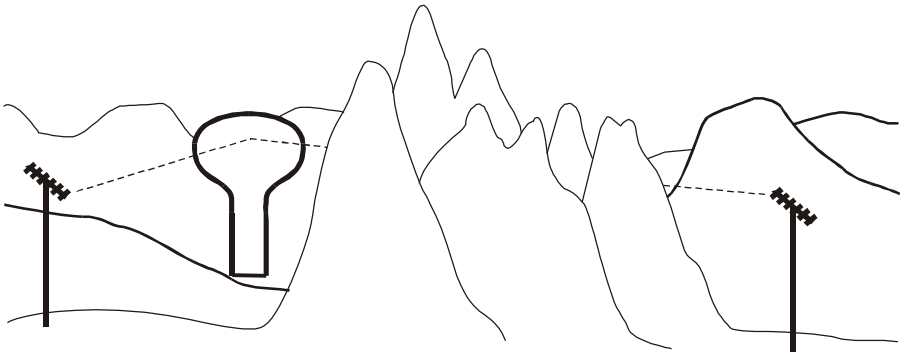

Figure 3-7. Reflection off Spherical Water Tower.

Under certain conditions, spherical water towers and walls of buildings (facing the proper direction) may enable communication around interfering terrain or vegetative obstacles (see fig. 3-7).

Transmitting over water allows VHF to go farther, but fading may occur. If communicating over water is unavoidable, and fading occurs, the signal may be improved by raising or lowering the antenna. The antenna may also be positioned so a hill or rise is between it and the water but not high enough to block the LOS to the other antenna (see fig. 3-8).

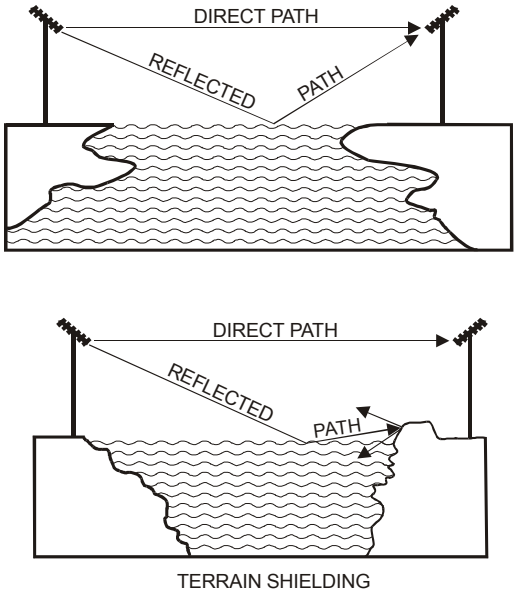

Figure 3-8. Multipath Fading and Terrain Shielding.

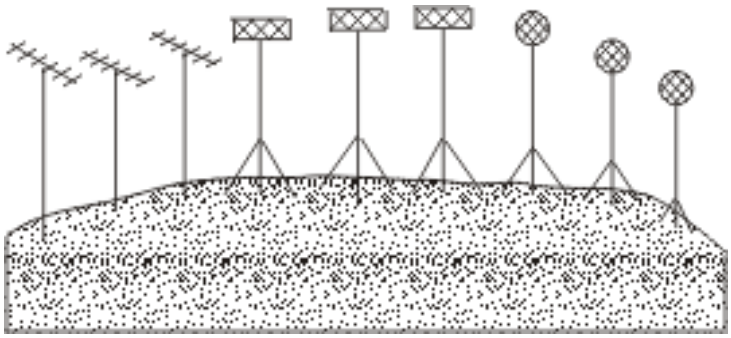

Figure 3-9. Ridge Line Antenna Farm.

Nothing is more compromising to a unit's location than an antenna farm stretched along a ridge line (see fig. 3-9). The enemy will realize that a major command post is nearby. The advantage of placing an antenna on a ridge line is the ability to talk in many directions without land being in the way. If communication is needed in only one direction—away from the enemy—put some terrain shielding between the antenna farm and the enemy (see figures 3-10 and 3-11). This way, the enemy won't be able to intercept communications or jam circuits as easily. However, it's not always necessary to talk from hilltop to hilltop. Talking from hillside to hillside or along the valley floor may be a better option in some instances. The enemy will certainly have a harder time locating a unit this way.

Figure 3-10. Low Antenna Profile.

Figure 3-11. Terrain Shielding.

The body can act as an antenna and affect the quality of the radio signal, particularly backpack and hand-held VHF sets with short antennas. The effect of the body on signal strength depends on frequency, antenna length, and the position of the antenna or set relative to the operator's body (see fig. 3-12 on page 3-8).

Backpack Sets (AN/PRC-119)

Maximum radiation (i.e., best performance) is to the front when the set is on the operator's back with a 3-foot whip antenna. The operator should then try facing in the direction of distant communications. This effect is most noticeable at frequencies greater than 50 MHz.

When the set is on the ground, and the operator is very close to the set, maximum radiation will probably occur through the operator's body. If the operator is a couple of feet away, the operator may act as a reflector and either improve or interfere with the signal.

Hand-Held Sets (AN/PRC-68)

The antenna of the hand-held AN/PRC-68 is much smaller than the AN/PRC-119's, and the operator's body affects directional characteristics to

Figure 3-12. Direction of Best Communications.

a greater extent, particularly when the antenna is lowered. This radio will normally be in the front jacket pocket, and best performance is then over the back. The higher frequencies are strongly affected when the antenna is lowered. Holding the radio in hand a few inches away from the body will modify the radiation pattern and can substantially lower performance to the sides at the higher frequencies. The directional characteristics of the antenna-body combination can be used to some advantage in reducing interference arriving from directions other than that of the signal. Trial and error is necessary to make this judgment.

VHF Siting

Position the antenna to reflect the directive pattern away from the wall or fence in the intended direction of communication when using VHF antennas near a metal-walled building or high, metal fence. For frequencies between 30 and 50 MHz, the antenna should be placed approximately 2 meters (6.5 feet) away; between 50 and 88 MHz, no more than 1 meter (3.2 feet) away. Communications may be reduced if antennas are placed more than 2 meters from the radio. The distance may be varied a

foot or so in each direction, while receiving, to find the position where the signal is the strongest when setting up the equipment. If frequencies are changed later, the optimum position will have to be redetermined at that time by the same method.

UHF Siting

At UHF and (to a much lesser degree) VHF frequencies, if an operator cannot see a person, then they probably cannot communicate—especially in heavy vegetation. Often, the signal has to travel up to the tops of the trees and move along the treetops and down to the receiver (see fig. 3-13). This will weaken it considerably. In vegetation, the antenna should be positioned away from trees that are in the direction of the signal and erected as high as possible. Changing the location of the antenna is also an option. Both horizontal and vertical orientations may be used with the AS-4225 Parabolic Grid Antenna with UHF multi-channel (MUX) radio equipment. Horizontal polarization is usually better for passing through the trees, but if the signal is skimming over the tops of the trees (probably unnoticeable) or over water, then the polarization should be vertical (see fig. 3-14 on page 3-10).

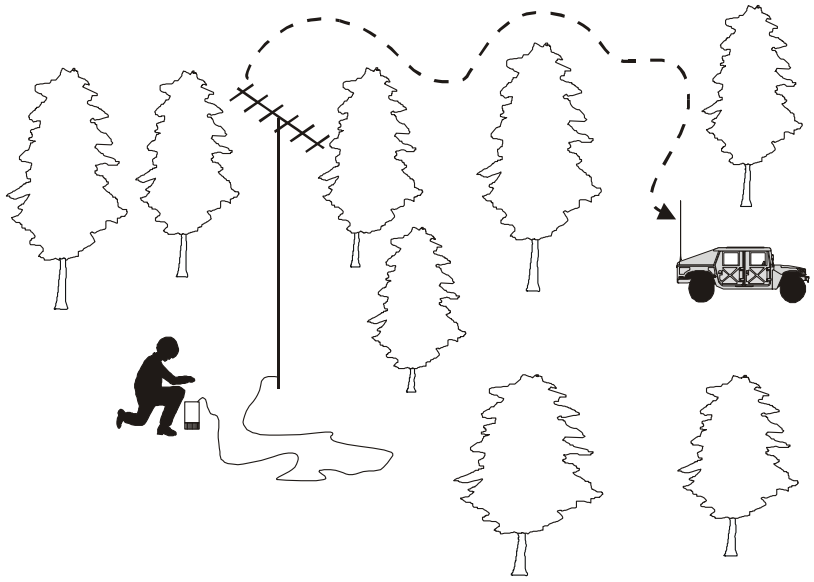

Figure 3-13. VHF Vegetative Propagation.

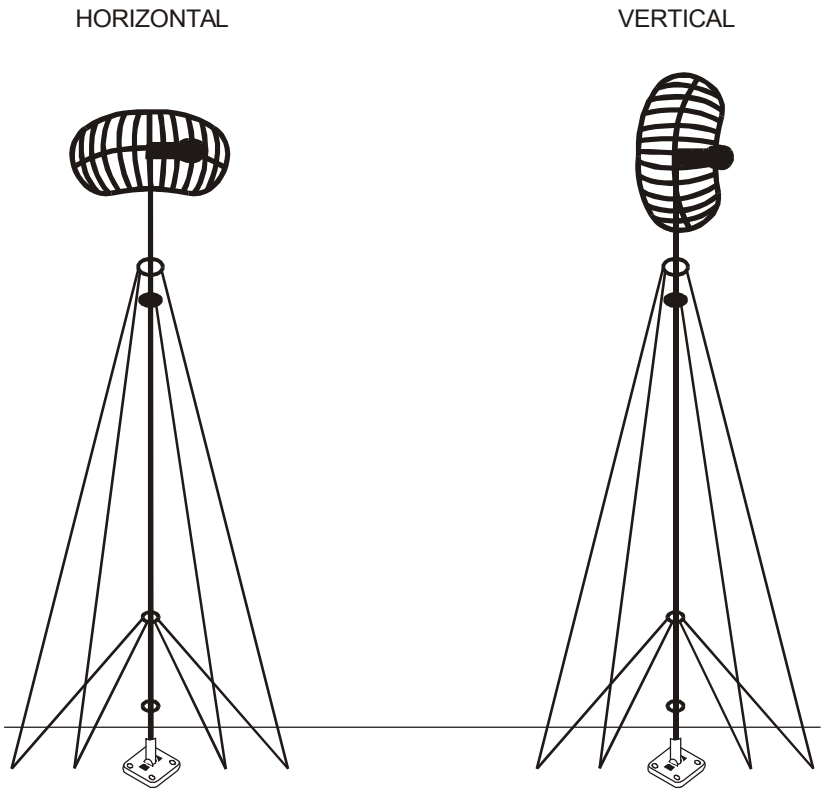

Figure 3-14. AS-4225 with AB-1356 Antenna Polarization.

GROUNDING TECHNIQUES

Poor grounding is probably the most important cause of a weak HF signal. Communication distances can easily be cut in half by improper grounding of the antenna. More importantly, the hazards involved with improper grounding coupled with high transmitter powers are bad burns, electrical shocks, or even death (see fig. 3-15).

Grounding prevents electrical shock to operator and improves signal strength, particularly in HF.

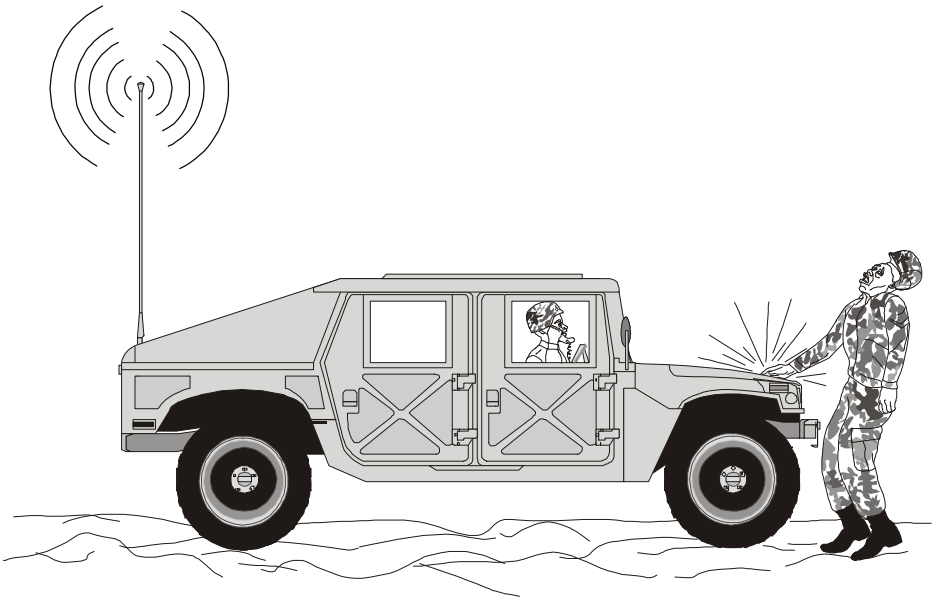

Figure 3-15. Shock Hazards of Ungrounded Equipment.

Ground Stake

The ground stake provided with the antenna should be driven deeply into the soil, making certain all connections are tight and clean. Soil moisture and salinity around the ground stake are very important for good grounding. If a dry or a damp location is available, choose the damp spot. If everything is bone dry, a couple of gallons of water poured around the stake may help. Adding a pound or two of salt from the mess tent to the soil around the stake before soaking it may help even more.

If a regulation ground stake is not available, there are many field-expedient means for grounding. The primary concern is to provide an electrical path from the equipment case, using braided copper or heavy gauge wire, to a buried metallic object that is in good contact with the ground. All cable connections and grounds should be free of grease, paint, or rust. Cables should be as short as possible.

Some useful grounds are—

- Metal fence posts.
- Steel reinforcing rods.

- Metal pipes.
- Metal plumbing (must not be connected to flammable liquid or gas).
- Metal building frames.

Ground Radials

A ground radial system (i.e., counterpoise) is necessary to reduce the amount of power lost in the earth (see fig. 3-16). This is particularly important for HF whips, inverted As, and other vertical antennas. The radial system design is usually a compromise between performance, portability, and time to install the system. Ground radials help to establish a known reference point of electrical ground. Without them, electrical ground may be some distance beneath the Earth's surface. Known electrical ground is important not only for formation of the wavefront off the antenna, but it also affects launch angles from antennas.

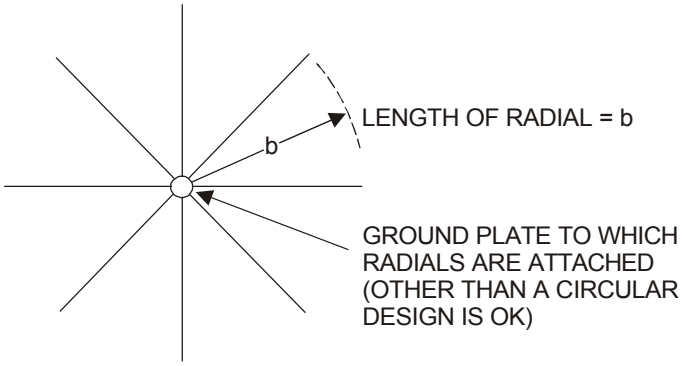

Figure 3-16. Radials.

Radials are attached metal-to-metal to a central point (a metal plate is often convenient). The radio frequency (RF) ground is attached to the central plate (Figure 3-17).

Wire diameter is not critical; select a diameter small enough to be lightweight and transportable, but large enough to prevent breakage. The largest number of radials to transport should be consistent with weight and bulkiness limitations. It is not necessary to make them greater than one-quarter wavelength at the lowest operating frequency.

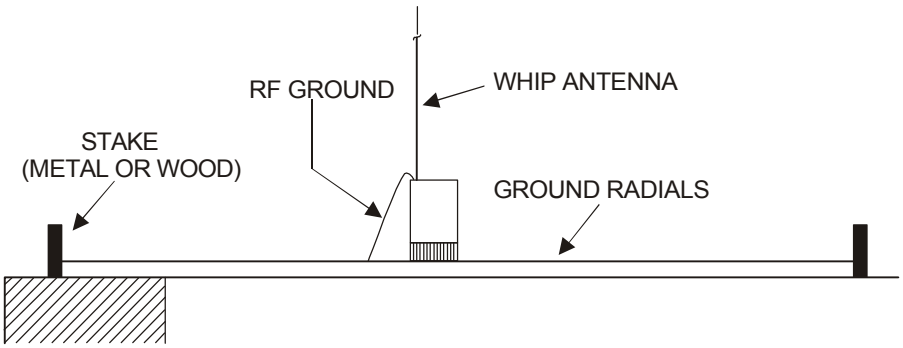

Figure 3-17. Equipment Configuration.

One-quarter wavelength = 74.9 meters
frequency (MHz)

Suggested dimensions:

5 meters (15-foot whip): N (number of radials) = at least 30 (one every 12°)
b (length of radial) = 7 meters (23 feet)

10 meters (32-foot whip): N = at least 30
b = 14 meters (46 feet)

If the site is semipermanent (i.e., several days or more), the number of radials should be increased to 100, and their length doubled.

Make radial systems for each antenna prior to an operation. Short stakes at the ends may be used to hold them in position when deploying. Ground all radios, where possible.

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Chapter 4

Antennas

This chapter will discuss high frequency and very high frequency antennas. Antenna designs that work for HF sometimes work for VHF and vice versa. The main difference is size; HF antennas are considerably larger than VHF antennas. Ultrahigh frequency antennas are generally limited to whips. When determining the best antenna to employ with a circuit, consult MCRP 6-22D for more detailed information on antenna propagation characteristics, as well as construction considerations and procedures.

HIGH FREQUENCY ANTENNAS

Vertical Whip

A vertical whip antenna (i.e., whip) will most likely be used with an HF radio. The whip is particularly good for ground wave communications in many directions at one time, at distances of 20 to 30 kilometers. Unfortunately, while it is radiating in all directions at the same time, it is also picking up interference from all directions. It is useless if using sky wave over a distance of 100 kilometers, because of the high radiation angles required. (NOTE: Launch angles off vertically polarized whip antennas are maximum below 45°.) A vertical whip's performance by sky wave, however, improves with increased path distances.

A vehicle-mounted whip, tied down fairly close to the vehicle, may be efficiently employed in short- and intermediate-distance HF sky wave communications. A whip, tied back only a little, may be useful in long-distance HF sky wave applications.

Sloping Wire

If an HF circuit is only a single point-to-point ground link or a ground wave net with all other terminals being located in the same direction, a

sloping wire may be used, if available. The radiating length of the AT-984 "Fishreel" antenna (a 45-foot long, wire antenna which can be used with the AN/PRC-104) can be varied by either connecting or disconnecting the alligator clips. (Antenna length is measured from the radio equipment.) See Table 4-1 to determine how long an antenna should be cut to form an assigned frequency.

Table 4-1. Sloping Wire Antenna Lengths.

Frequency	Length	
MHz	(Feet)	(Meters)
5	45	13.7
9	25	7.6
12	15	4.6

The far end of the antenna should be connected to a rope with a weight, such as a stone or brick or other nonconducting material tied to the end. The weighted end should be thrown over a tree so that the antenna forms a 30° to 45° angle to the ground. Angles higher than 45° should be used for ground wave, and lower angles for sky wave. The high end should be opposite the direction of the intended receiver (see fig. 4-1).

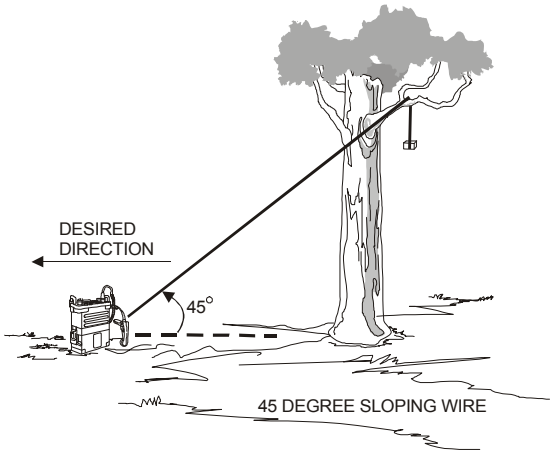

Figure 4-1. 45° Sloping Wire Antenna.

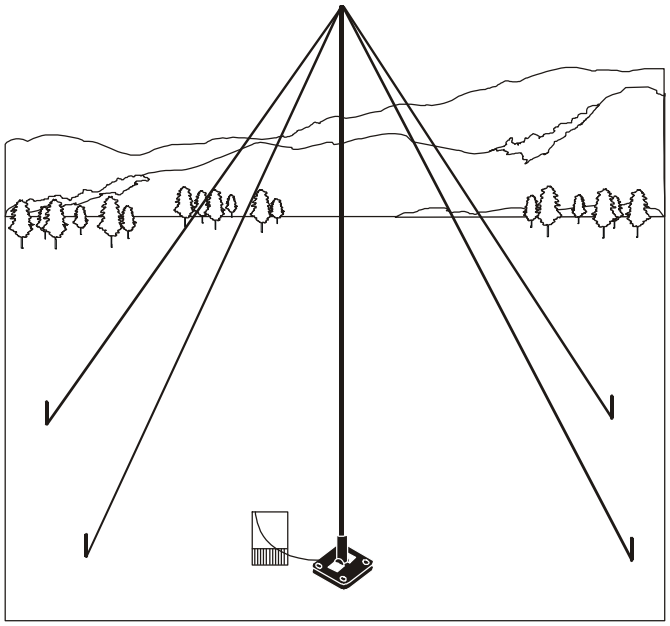

Figure 4-2. AS-2259.

AS-2259 Near-Vertical Incidence

Another HF antenna is the AS-2259 (see fig. 4-2). Although not suitable for frequencies under 3.5 MHz, this antenna can be used for both ground wave and sky wave. The AS-2259 will often enable an operator to communicate in a skip zone when a whip antenna will not. In addition, it can sometimes be used effectively to communicate by sky wave over a hill or mountain obstacle that would otherwise block a ground wave signal. More detailed information regarding HF NVIS antennas can be found in MCRP 6-22D.

Horizontal Half -Wave Dipole (Doublet)

The horizontal half-wave dipole (also known as the doublet) is frequently used for short to medium HF sky wave paths (up to about 1,500 kilometers). It is usually installed at one-quarter wavelength of the operating frequency above ground (see fig. 4-3 on page 4-4). The major drawback of this antenna is the unusually long length required (up to 71

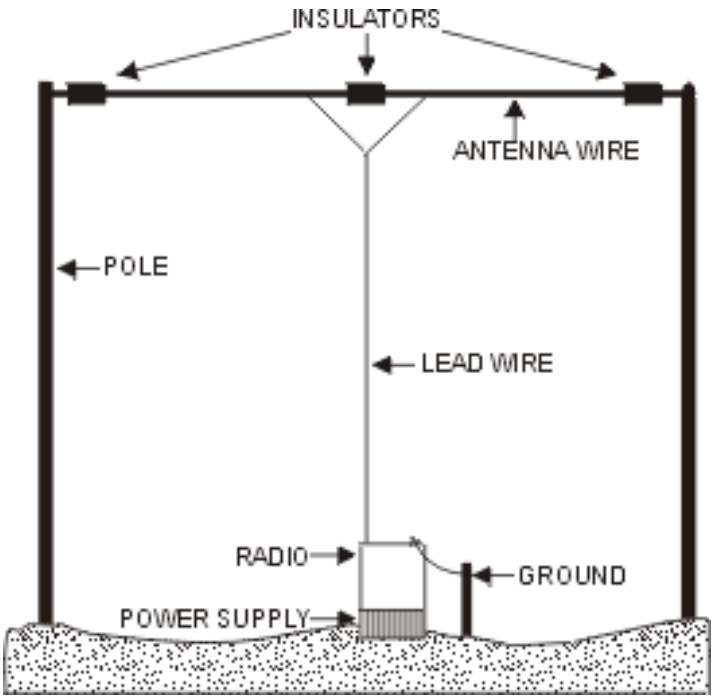

Figure 4-3. Doublet Antenna.

meters [233 feet] at 2 MHz). NOTE: A 2 percent or greater error in length means less efficiency and a loss of radiated power.

Inverted L

Inverted L antennas are useful for NVIS propagation (see fig. 4-4). Because of their construction, they also yield better ground wave radiation than whip antennas. Inverted L antennas can be effectively employed with a wider range of frequencies than can horizontal half-wave dipole antennas.

Antenna Enhancements

Several pieces of equipment which improve the capabilities of standard Marine Corps communications equipment are available from commercial sources. These items are not stocked in the Marine Corps supply system. They must be purchased directly from commercial sources.

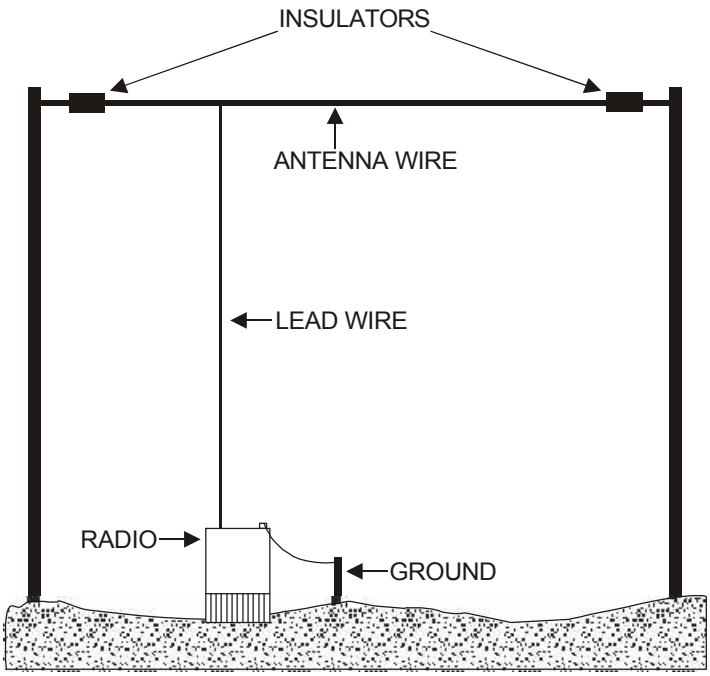

Figure 4-4. Inverted L.

They include—

- Tilt whip adapter (TWA)—used with the vertical whip antennas issued with AN/MRC-138 radio. These antennas can be tilted to obtain the correct angles for NVIS communications. The TWA made for the AN/MRC-138 is one piece and has a simple design. When fitted with a TWA and only the top four sections of the AT-1011 antenna, the AN/MRC-138 can be operated on the move.
- Whip loading coil (WLC)—used with the WTA for the AN/MRC-138. This device makes the antenna more efficient at lower frequencies.
- Whip to wire adapter (WWA)—screws into the top one-inch-diameter section of the AT-1011, thus allowing the operator to use the one-inch-diameter section of the AT-1011 as the mast for field-expedient antennas.

- Whip-base adapter (WBA)—allows the whip antennas for Army HF radios to be used with Marine Corps HF radios. The whip antennas for Army HF radios are of a different size than the whip antennas for Marine Corps HF radios.

For further information about commercial equipment which can improve the capabilities of Marine Corps radios, contact the G-6/S-6 sections of higher headquarters.

VERY HIGH FREQUENCY ANTENNAS

OE-254

This antenna is used with VHF-FM radios to increase the operating range beyond that of a normal whip. The two elements hanging down from the antenna form a ground-plane similar to the ground radials discussed in chapter 3. The effect is to act as an artificial ground and greatly increase the signal range. The antenna radiates in all directions at the same time. The OE-254's elements allow it to tune to frequencies between 30 and 88 MHz without manually adjusting either the ground-plane or radiating elements' length. Figure 4-5 illustrates the OE-254 antenna.

Vertical Whip

VHF whip antennas are usually limited in range from 15 to 20 miles. A whip antenna is omnidirectional; therefore, it has the potential to produce a great deal of radio-wave interference with the radios in the area. It is important when using a whip antenna or any antenna to keep maximum distance between antennas. Use of hills and other terrain features to block off unwanted signals will improve desired signal strength. When using whip antennas with backpack or hand-held equipment, body position may increase the transmitted and received signal. This is because the human body acts as an antenna. If experiencing communication problems, the operator should try facing in different directions to improve the reception of the signal.

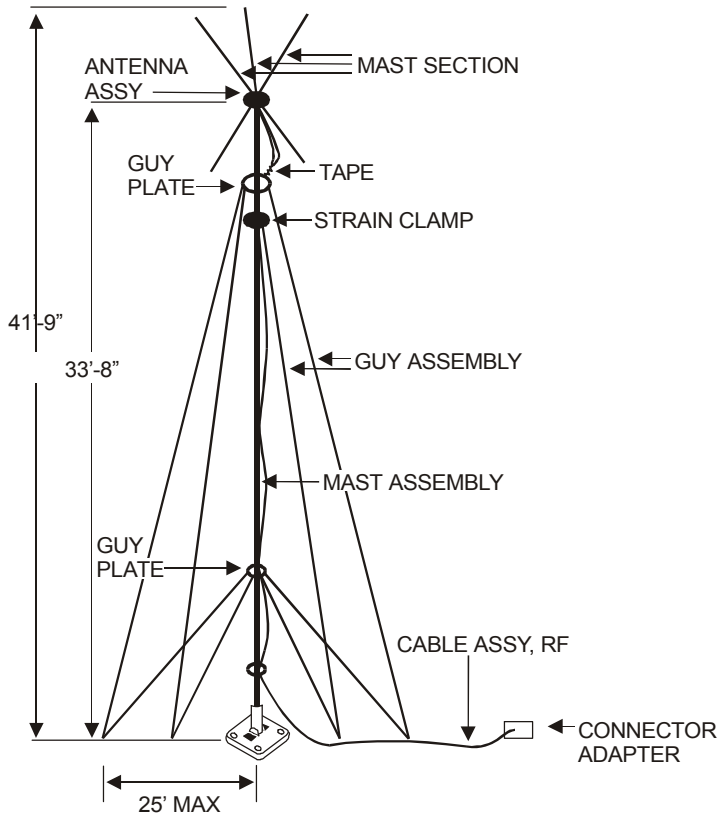

Figure 4-5. OE-254 Antenna.

ANTENNA LENGTH

The length of an antenna must be considered in two ways. It has both a physical and an electrical length, and the two are never the same. The reduced velocity of the wave on the antenna and a capacitive effect (known as end effect) make the antenna seem longer electrically than it is physically. The contributing factors are the ratio of the diameter of the antenna to its length and the capacitive effect of terminal equipment (e.g., insulators, clamps, etc.) used to support the antenna.

To calculate the physical length of an antenna, use a correction of 0.95 for frequencies between 3.0 and 50.0 MHz. The following figures are for a half-wave antenna.

Length (meters) = 150 x 0.95/Frequency in MHz
= 142.5/Frequency in MHz

Length (feet) = 492 x 0.95/Frequency in MHz
= 468/Frequency in MHz

The length of a long-wire antenna (one wavelength or longer) for harmonic operation is calculated by using the following formula:

Length (meters) = 150(N-0.05)/Frequency in MHz
Length (feet) = 492(N-0.05)/Frequency in MHz

N = number of half-wave lengths in the total length of the antenna.

For example, if the number of half-wavelengths is 3 and the frequency in MHz is 7, then:

Length (meters) = 150(N-0.05)/Frequency in MHz
= 150(3-.05)/7
= 150 x 2.95/7
= 442.50/7
= 63.2 meters

Chapter 5

Interference

Radio frequency interference is always present in a military environment. It may come from a single source or a combination of many sources including natural or manmade frequency interference, poor equipment condition, improper equipment usage, frequency interference, use of unauthorized frequencies, and frequency reuse.

NATURAL INTERFERENCE

Natural radio noise has two principal sources: thunderstorms (atmospheric noise) and stars (galactic noise). It is especially noticeable at night when the lower frequencies propagate farther than in the daytime. The only way to reduce this type of interference is to use a directional antenna to prevent receiving the interference from all directions. However, this will not eliminate the noise coming from the direction of the received signal. Use of a higher frequency will also help, although if a sky wave circuit is used, care must be exercised not to pick the highest frequency at which the signal will be refracted to Earth by the ionosphere (i.e., the critical frequency).

MANMADE INTERFERENCE

Most manmade interference comes from electrical sources such as power generators, alarm systems, power lines, auto ignition, fluorescent lighting, faulty electrical relay contacts, and electrified railroads. Manmade interference also includes enemy jammers (see chapter 7). The key to combating this form of interference is to isolate communications equipment from manmade interference. The interference from known sources such as generators can be greatly reduced if an antenna is positioned so that an obstacle (e.g., a hill) is between it and the source. This must be done so that the same obstacle will not block the intended radio path. If the interference is not coming from the same direction as the intended signal, then a directional antenna should be used.

POOR EQUIPMENT CONDITION AND IMPROPER USAGE

The condition of radio equipment and how it is being used may result in interference. There are several steps that should be taken to lessen this possibility. These include making certain that shielded cables are used where required, ensuring connectors are properly connected to cables, and making sure that antennas within a group are as far apart as possible. All antenna leads (transmission lines), power lines, and telephone lines should be as short as possible when they are on the ground and should not cross. If lines do cross, they must cross at 90° angles to each other, and they must be separated from each other by standoffs. Lines threaded through the trees near an antenna serve as pipelines for interference to and from antennas. Finally, ensure that all radio equipment is grounded.

FREQUENCY INTERFERENCE AND INTERMODULATION

Frequency interference is one of the easiest communications problems to prevent, but it must be done by the CIS officer, CIS chief, and frequency manager during the development of the CIS plan. This type of interference is caused primarily by two radio transmitters using the same frequency, but it can also happen when different frequencies are used. Most of these problems can be eliminated by good frequency planning. However, if frequency interference does occur the following steps can be taken to improve communications:

- Identify the source of the interference. If it is a VHF or HF ground wave transmission, it will probably be within the immediate area and will only occur when the offending transmitter is keyed. The other operator is probably transmitting on a different frequency and has no way of knowing that he is interfering with anyone else's ability to transmit.
- Get the interfering operator to lower their transmitter power as long as it does not degrade their circuit.
- Put as much distance as possible between the affected unit's equipment. This may involve using a hill or other object to block the signal.
- Change to a directional antenna.

- Remember that the receiver of interference may also cause someone else interference and, whenever possible, should lower his power and use a directional antenna.
- Report interference to the CIS officer or CIS chief.

USE OF UNAUTHORIZED FREQUENCIES

There is one final source of frequency interference: the use of unauthorized frequencies. This practice is illegal and has the potential to disrupt a carefully engineered frequency plan, introduce interference to other frequencies and circuits, and prevent other units from fulfilling their mission. Radio operators should never use unauthorized frequencies.

FREQUENCY REUSE

There are not enough radio frequencies available for all radio operators to have their own channel. When HF propagation conditions are favorable, Marines may discover that their radio frequency is being used by foreign or United States military personnel in other countries. VHF FM frequencies often have to be reused within the same operation by more than one unit. The exercise frequency manager will try to make certain that users of the same frequency are as far away as possible from each other, but some units (United States Marine Corps and Army, in particular) will join at some stage in the operation. When this occurs, the first common, higher headquarters should be informed to settle the problem.

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Chapter 6

Radio Operations Under Unusual Conditions

OPERATIONS IN DESERT AREAS

Capabilities and Limitations

SCR is usually the primary means of communications in the desert. It can be employed effectively in desert climate and terrain to provide the highly mobile means of communications demanded by widely dispersed forces. However, desert terrain provides poor electrical ground, and counterpoises are needed to improve operation.

Techniques for Operations

For the best operation in the desert, radio antennas should be located on the highest terrain available. Transmitters using whip antennas in the desert will lose one-fifth to one-third of their normal range because of the poor electrical grounding characteristics of desert terrain. For this reason, it is important to use complete antenna systems such as horizontal dipoles and vertical antennas with adequate counterpoises.

Equipment Considerations

Some SCRs automatically switch on their second blower fan if their internal temperature rises too high. Normally, this happens only in temperate climates when the radios are transmitting. This may disturb Marines unaccustomed to radio operation in the desert environment. Operation of the second fan, however, is quite normal. RF power amplifiers used in AM and SSB sets are liable to overheat severely and burn out. They should be turned on only when necessary (signal reception is not affected). Since the RF power amplifiers take approximately 90 seconds to reach the operating mode, the standing operating procedure (SOP) of units using the equipment should allow for delays in replying.

Dust affects communications equipment such as SSB/AM RF power amplifiers. Dust covers should be used whenever possible. Some receiver-transmitter units have ventilating ports and channels that can get clogged with dust. These must be checked regularly and kept clean to prevent overheating.

Batteries. Wet cell batteries do not hold their charge efficiently in intense heat. Electrolyte evaporates rapidly and should be checked weekly (more often, if warranted). Add distilled water as needed. Extra containers of distilled water should be carried in the vehicle. Maintenance of vehicle batteries, beyond adding water, must be done only by authorized motor-transport personnel according to applicable Marine Corps Orders and SOPs. Dry battery supplies must be increased, since hot weather causes batteries to fail more rapidly.

Electrical Insulation. Wind-blown sand and grit will damage electrical wire insulation over a period of time. All cables that are likely to be damaged should be protected with tape before insulation becomes worn. Sand will also find its way into parts of items such as "spaghetti cord" plugs, either preventing electrical contact or making it impossible to join the plugs. Carry a brush, such as an old toothbrush, and use it to clean such items before they are joined.

Condensation. In deserts with relatively high dew levels and high humidity, overnight condensation can occur wherever surfaces (such as metals exposed to air) are cooler than the air temperature. This condensation can affect electrical plugs, jacks, and connectors. All connectors likely to be affected by condensation should be taped to prevent moisture from contaminating the contacts. Plugs should be dried before inserting them into equipment jacks. Excessive moisture or dew should be dried from antenna connectors to prevent arcing.

Static Electricity. Static electricity is prevalent in the desert. It is caused by many factors, e.g., wind-blown dust particles. Extremely low humidity contributes to static discharges between charged particles. Poor grounding conditions exacerbate the problem. Be sure to tape all sharp edges (tips) of antennas to cut down on wind-caused static discharges and the accompanying noise. If you are operating from a fixed position,

ensure that equipment is properly grounded. Since static-caused noise diminishes with an increase in frequency, use the highest frequencies that are available and authorized.

Maintenance Improvement

In desert areas, the maintenance of SCRs becomes more difficult because of the large amounts of sand, dust, or dirt that enter the equipment. Radios equipped with servomechanisms are particularly affected. To reduce maintenance downtime, keep the radios in dustproof containers as much as possible. It is also important to keep air vent filters clean to allow cool air to circulate to prevent overheating. Preventive maintenance checks should be made frequently. Also, keep a close check on lubricated parts of the equipment. If dust and dirt mix with the lubricants, moving parts may be damaged.

OPERATIONS IN JUNGLE AREAS

Capabilities and Limitations

SCR communications in jungle areas must be carefully planned because the dense jungle growth significantly reduces the range of radio transmission. However, since SCR can be deployed in many configurations, especially manpacked, it is a valuable communications asset. Mobility is also an advantage of SCR. The capabilities and limitations of SCR must be carefully considered when used by forces in a jungle environment.

Climate and density of jungle growth limits SCR communications in jungle areas. The hot and humid climate increases the maintenance problems of keeping equipment operable. Thick jungle growth acts as a vertically polarized absorbing screen for RF energy that reduces transmission range. Therefore, increased emphasis on maintenance and antenna siting is necessary when operating in jungle areas.

Techniques for Operations

The main problem in establishing SCR communications in jungle areas is the siting of the antenna.

Apply the following techniques to improve communications in the jungle:

- Antennas should be located in clearings on the edge farthest from the distant station and as high as possible.
- Antenna cables and connectors should be kept off the ground to lessen the effects of moisture, fungus, and insects. This also applies to all power and telephone cables.
- Complete antenna systems, such as ground planes and dipoles, are more effective than fractional wavelength whip antennas.
- Vegetation must be cleared from antenna sites. If an antenna touches any foliage, especially wet foliage, the signal will be grounded.
- Vegetation, particularly when wet, acts like a vertically polarized screen and absorbs much of a vertically polarized signal. Horizontally polarized antennas are preferred to vertically polarized antennas.

Maintenance Improvement

Because of moisture and fungus, the maintenance of SCR in tropical climates is more difficult than in temperate climates. The high relative humidity causes condensation to form on the equipment and encourages the growth of fungus. Operators and maintenance personnel should check the appropriate TMs for any special maintenance requirements. Some techniques for improving maintenance in jungle areas are—

- Keep the equipment as dry as possible and in lighted areas to retard fungus growth.
- Keep all air vents clear of obstructions so air can circulate to cool and dry the equipment.
- Keep connectors, cables, and bare metal parts as free of fungus growth as possible. Use moisture fungusproofing paint (MFP) to protect equipment after repairs are made or when equipment is damaged or scratched.

High Frequency Expedient Antennas

Dismounted patrols and units of company size and below can greatly improve their ability to communicate in the jungle by using expedient antennas. While moving, they are generally restricted to using the short and long antennas which come with the radios. However, when they are not moving, these expedient antennas will allow them to broadcast farther and receive more clearly.

Note: An antenna that is not "tuned" or "cut" to the operating frequency is not as effective as the whips that are supplied with the radio.

Circuits inside the radio "load" the whips properly so that they are "tuned" to give maximum output. Whips are not as effective as a tuned doublet or tuned ground-plane, but the doublet or ground-plane must be tuned to the operating frequency.

OPERATIONS IN A COLD WEATHER ENVIRONMENT

Capabilities and Limitations

SCR equipment has certain capabilities and limitations that must be carefully considered when operating in extremely cold areas. However, in spite of significant limitations, SCR is the normal means of communications in such areas.

One of the most important capabilities of SCR in cold weather areas is its versatility. Vehicular-mounted radios can be moved relatively easily to almost any point where it is possible to install a command headquarters. Smaller, manpacked radios can be carried to any point accessible by foot or aircraft.

A limitation on radio communications that radio operators must expect in extremely cold areas is interference by ionospheric disturbances. These disturbances, known as ionospheric storms, have a definite degrading effect on sky wave propagation. Moreover, either the storms or the auroral (e.g., Northern Lights) activity can cause complete failure of radio communications. Some frequencies may be blocked out completely by static for extended periods of time during storm activity.

Fading, caused by changes in the density and height of the ionosphere, can also occur and may last from minutes to weeks. The occurrence of these disturbances is difficult to predict. When they occur, the use of alternate frequencies and a greater reliance on FM or other means of communications are required.

Techniques for Operations

Whenever possible, SCR for tactical operations in cold weather areas should be installed in vehicles to reduce the problem of transportation and shelter for operators. This will also help solve some of the grounding and antenna installation problems caused by the climate.

Because of permafrost and deep snow, it is difficult to establish good electrical grounding in extremely cold areas. The conductivity of frozen ground is often too low to provide good ground wave propagation. To improve ground wave operation, use a counterpoise to offset the degrading effects of poor electrical ground conductivity. Remember to install a counterpoise high enough above the ground so that it will not be covered by snow.

In general, antenna installation in arcticlike areas presents no serious difficulties. However, installing some antennas may take longer because of adverse working conditions. A few tips for installing antennas in extremely cold areas are—

- Mast sections and antenna cables must be handled carefully since they become brittle in very low temperatures.
- Antenna cables should be constructed overhead to prevent damage from heavy snow and frost, whenever possible. Nylon rope guys, if available, should be used in preference to cotton or hemp because nylon ropes do not readily absorb moisture and are less likely to freeze and break.
- Antennas should have extra guy wires, supports, and anchor stakes to withstand heavy ice and wind loading.

Some Marine Corps radios that are adjusted to a particular frequency in a relatively warm place may drift off frequency when exposed to

extreme cold. Low battery voltage can also cause frequency drift. When possible, allow a radio to warm up several minutes before placing it into operation. Since extreme cold tends to lower output voltage of a dry battery, try warming the battery with body heat before operating the radio set. This minimizes frequency drift. Flakes or pellets of highly electrically charged snow are sometimes experienced in northern regions. When these particles strike the antenna, the resulting electrical discharge causes a high-pitched static roar that can blanket all frequencies. To overcome this static, antenna elements can be covered with polystyrene tape and shellac.

Maintenance Improvement

The maintenance of SCR equipment in extreme cold presents many difficulties. Radios must be protected from blowing snow, since snow will freeze to dials and knobs and blow into the wiring to cause shorts and grounds. Cords and cables must be handled carefully since they may lose their flexibility in extreme cold. All radio equipment and power units must be properly winterized. Check the appropriate technical manual (TM) for winterization procedures. A few tips for maintenance in arctic areas are discussed in the following paragraphs.

Power Units. As the temperature goes down, it becomes increasingly difficult to operate and maintain generators. They should be protected as much as possible from the weather.

Batteries. The effect of cold weather on wet and dry cell batteries depends upon the type and kind of battery, the load on the battery, the particular use of the battery, and the degree of exposure to cold temperatures.

Shock Damage. Damage may occur to vehicular SCR by the jolting of the vehicle. Most synthetic rubber shock mounts become stiff and brittle in extreme cold and fail to cushion equipment. Check the shock mounts frequently and change them, as required.

Winterization. Check the TMs for the SCR and power source to see if there are special precautions for operation in extremely cold climates.

For example, normal lubricants may solidify and permit damage or malfunctions. They must be replaced with the recommended arctic lubricants.

Microphones. Moisture from your breath may freeze on the perforated cover plate of your microphone. Use standard microphone covers to prevent this. If standard covers are not available, improvise a suitable cover from rubber or cellophane membranes or from rayon or nylon cloth.

Breathing and Sweating. A SCR generates heat when it is operated. When it is turned off, the air inside cools and contracts and draws cold air into the set from the outside. This is called breathing. When a radio breathes, and the still-hot parts come in contact with subzero air, the glass, plastic, and ceramic parts of the set may cool too rapidly and break.

When cold equipment is brought suddenly into contact with warm air, moisture will condense on the equipment parts. This is called sweating. Before cold equipment is brought into a heated area, it should be wrapped in a blanket or parka to ensure that it will warm gradually to reduce sweating. Equipment must be thoroughly dry before it is taken back out into the cold air or the moisture will freeze.

Vehicular-Mounted Radios. These radios present special problems during winter operations because of their continuous exposure to the elements. Proper starting procedures must be observed. The radio's power switch must be off prior to starting the vehicle; this is a particularly critical requirement when vehicles are slave started. If the radio is cold soaked from prolonged shutdown, frost may have collected inside the radio and could cause circuit arcing. Hence, time should be allowed for the vehicle heater to warm the radio sufficiently that any frost collected within the radio has a chance to thaw. This may take up to an hour. Once the radio has been turned on, it should warm up for approximately 15 minutes before transmitting or changing frequencies. This allows components to stabilize. If a vehicle is operated at a low idle with radios, heater, and lights on, the batteries may run down. Before increasing engine revolutions per minute to charge the batteries, radios should be turned off to avoid an excessive power surge. A light coat of silicon compound on

antenna mast connections helps to keep them from freezing together and becoming hard to dismantle.

OPERATIONS IN MOUNTAINOUS AREAS

Capabilities and Limitations

Operation of SCRs in mountainous areas has many of the same problems as in northern or cold weather areas. Also, the mountainous terrain makes the selection of transmission sites a critical task. In addition, the terrain restrictions encountered frequently make radio relay stations necessary for good communications.

Maintenance Improvement

Because of terrain obstacles, SCR transmissions will frequently have to be by line of sight. Also, the ground in mountainous areas is often a poor electrical conductor. Thus, a complete antenna system, such as a dipole or ground-plane antenna with a counterpoise, should be used. The maintenance procedures required in mountainous areas are very often the same as maintenance in northern or cold weather areas. The varied or seasonal temperature and climatic conditions in mountainous areas make flexible maintenance planning a necessity.

OPERATIONS IN SPECIAL ENVIRONMENTS

Urbanized Terrain

SCR communications in urbanized terrain poses special problems. Some problems are similar to those encountered in mountainous areas. There are problems of obstacles blocking transmission paths. There is the problem of poor electrical conductivity because of pavement surfaces. There is also the problem of commercial power-line interference. VHF radios are not as effective in urbanized terrain as they are in some other areas. The power output and operating frequencies of these VHF radios require a line of sight between antennas. Line of sight at street level is not always possible in built-up areas.

HF radios do not require or rely on line of sight as much as VHF radios because operating frequencies are lower, and power output is greater. The problem is that HF radios are not organic to small units. To overcome this, the VHF signals must be retransmitted.

Retransmission stations in aerial platforms can provide the most effective means if they are available. Organic retransmission is more likely to be used. The antenna should be hidden or blended in with surroundings. This will help prevent the enemy from using it as a landmark to "home in" his artillery bombardment. Antennas can be concealed by water towers, existing civilian antennas, or steeples.

The following steps should also be taken within urbanized terrain:

- Park radio-equipped vehicles inside buildings for cover and concealment.
- Dismount radio equipment and install it inside buildings (in basement, if available).
- Place generators against buildings or under sheds to decrease noise and provide concealment (adequate ventilation must be provided to prevent heat buildup and subsequent failure of generator).

Nuclear, Biological, and Chemical Environment

One of the realities of fighting on the modern battlefield is the presence of nuclear weapons. The explosion of a nuclear weapon causes a tremendous blast, followed by intense heat and strong radiation. The ionization of the atmosphere by a nuclear explosion will have degrading effects on communications because of static and the disruption of the ionosphere.

Another effect of a nuclear explosion that is an even greater danger to radio communications is the electromagnetic pulse (EMP). EMP is a strong pulse of electromagnetic radiation, many times stronger than the static pulse generated by lightning. This pulse can enter the radio through the antenna system, power connections, and signal input connections. In the equipment, the pulse can break down circuit components such as transistors, diodes, and integrated circuits. It can melt capacitors, inductors, and transformers. EMP can destroy a radio.

Defensive measures against EMP call for proper maintenance, particularly the shielding of equipment. When the equipment is not in use, all antennas and cables should be removed to decrease the effect of EMP on the equipment. Effective grounding is necessary to reduce effect of EMP. EMP is a danger to SCR equipment, but contamination is a danger to Marines. Contamination from any portion of the nuclear, biological, and chemical (NBC) environment has adverse effects on both equipment and personnel.

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Chapter 7

Electronic Warfare

Electronic warfare (EW) is the military action involving the use of electromagnetic energy (i.e., radio frequency waves) to attack personnel, facilities, or equipment with the intent of degrading, neutralizing, or destroying enemy combat capability. EW includes electronic attack (EA), electronic protection (EP), and electronic warfare support (ES).

EA includes actions taken to prevent or reduce the enemy's effective use of the electromagnetic spectrum and employment of weapons that use electromagnetic or directed energy. EP represents actions taken to protect personnel, facilities, and equipment from any effects of friendly or enemy employment of electronic warfare that degrade, neutralize, or destroy friendly combat capability. ES involves actions taken by, or under direct control of, an operational commander to search for, intercept, identify, and locate sources of intentional or unintentional radiated electromagnetic energy for the purpose of immediate threat recognition. Each radio operator must be aware of what the enemy will try to do. The enemy is well equipped to conduct EW, and the different techniques the enemy uses have specific purposes in the enemy's EW effort.

ELECTRONIC ATTACK TECHNIQUES

Enemy forces employ a large number of radio direction finder (RDF) sets and communications intelligence (COMINT) analysts to exploit friendly use of the electromagnetic spectrum. The enemy's goal is to locate and destroy as many command and control, fire support, and intelligence sites as possible during the first critical phase of the battle. When the enemy locates sites that the enemy cannot or does not want to destroy, these sites become prime targets for imitative electronic deception (IED) or jamming. Imitative electronic deception is the enemy's use of a compatible radio and a language expert to enter a friendly radio net. The enemy IED experts are very good at their jobs. If they are permitted to enter into a net, they will create much confusion for friendly forces.

Jamming is an effective way to disrupt control of the battle. All it takes is a transmitter, tuned to your frequency, with the same type of modulation and with enough power output to override the signal at your receiver.

There are many types of jamming signals that may be used against a radio operator. Some are very difficult to detect and some are impossible to detect. For this reason, an operator must always be alert to the probability of jamming and react accordingly when the radio has been silent for an inordinate amount of time. The radio operator should also be able to quickly identify the various types of jamming signals. These include—

- Random noise.
- Random pulse.
- Stepped tones.
- Wobbler.
- Random keyed modulated continuous wave.
- Tone.
- Rotary.
- Pulse.
- Spark.
- Recorded Sounds.
- Gulls.
- Sweep-through.

Capture Effect and Jamming Techniques

An inherent characteristic in FM communications is that a given station transmitting a signal will capture those receivers on the same frequency and in range for the receiver to detect the signal. This is the basis for netted communications for VHF FM radios. This FM capture effect is undesirable when receivers in a net are "captured" by a transmitter not in that net. This could be friendly interference or enemy interference. Friendly interference is usually unintentional whereas enemy interference is usually intentional.

Obvious Jamming

Radio operators are mostly aware of obvious interference (e.g., jamming) by an enemy, such as stepped tones (e.g., bagpipes), random-keyed Morse Code, pulses, and recorded sounds. The purpose of this type of jamming is to block out reception of friendly transmitted signals and to cause a nuisance to the receiving operator. An operator can usually detect when the enemy is using this type of jamming.

Subtle Jamming

This type of jamming is not obvious at all. With subtle jamming, no sound is heard from the receiver. The radio does not receive incoming friendly signals, yet everything seems normal to the operator.

Operator Actions

Radio operators must be able to determine whether or not their radios are being jammed. This is not always easy. Threat jammers may employ obvious or subtle jamming techniques. These techniques may consist of powerful unmodulated or noise-modulated carrier signals transmitted to the operator's receiver. Unmodulated jamming signals are characterized by a lack of noise. Noise-modulated jamming signals are characterized by obvious interference noises. If radio operators suspect that their radios are the targets of threat jamming, the following procedures will help them to make this determination.

Meaconing, Intrusion, Jamming, and Interference Report

If the radio operator suspects jamming or enemy intrusion on the net, then the radio operator should report it immediately to higher headquarters. Such information is vital for the protection and defense of radio communications.

Field meaconing, intrusion, jamming, and interference (MIJI) reports serve two purposes. First, initial MIJI reports facilitate battlefield evaluations of the enemy's actions or intentions and provide data for tactical countermeasures, as appropriate. Second, complete and accurate follow-up reports ensure MIJI incidents are documented and evaluated on a

national level, thus providing data for a continuing study of foreign electronic warfare capabilities and activities.

MIJI reports may be transmitted over nonsecure electronic means when secure communications are not available; however, the textual content of the MIJI report will be secured by an off-line (i.e., manual) system. Reports will be prepared in the format outlined in the following paragraphs. Brevity numbers pertinent to specific line item information are provided for some items. These brevity numbers must be encoded in the numeral cipher or authentication system prior to transmission. The two types of field MIJI reports are—

- MIJI 1—an abbreviated initial report containing only those items of information necessary to inform headquarters of the incident and enable them to initiate evaluatory or retaliatory actions as appropriate.
- MIJI 2—consists of 40 lines and is completed by higher headquarters.

The MIJI 1 Report. This report is forwarded through the chain of command to the combat operations center by the operator who is experiencing the MIJI incident. A separate report is submitted for each MIJI incident. The MIJI report includes—

- Item 1—type report. When being transmitted over nonsecure communications means, the numerals 022 are encrypted as Item 1 of the MIJI 1 report. When being transmitted over secure communications means, the term MIJI 1 is used as Item 1 of the MIJI 1 report.
- Item 2—type MIJI incident. When being transmitted over nonsecure communications means, the appropriate numeral preceding one of the items below is encrypted as Item 2 of the MIJI 1 report. When being transmitted over secure communications means, the appropriate term below is used as Item 2 of the MIJI 1 report.

  1 Meaconing
  2 Intrusion
  3 Jamming
  4 Interference

- Item 3—type of equipment affected. When being transmitted over nonsecure communications means, the appropriate numeral preceding one of the terms below is encrypted as Item 3 of the MIJI 1 report. When being transmitted over secure communications means, the appropriate term below is used as Item 3 of the MIJI 1 report.

  1 Radio
  2 Radar
  3 Navigational aid
  4 Satellite
  5 Electro-optics

- Item 4—Frequency or channel affected. When being transmitted over nonsecure communications means, the frequency or channel affected by the MIJI incident is encrypted as Item 4 of the MIJI 1 report. When being transmitted over secure communications means, the frequency or channel affected by the MIJI incident is Item 4 of the MIJI 1 report.
- Item 5—victim designation and call sign of affected station operator. The complete call sign of the affected station operator is Item 5 of the MIJI 1 report over both secure and nonsecure communications means.
- Item 6—coordinates of the affected station. When being transmitted over nonsecure communications means, the complete grid coordinates of the affected station are encrypted as Item 6 of the MIJI 1 report. When being transmitted over secure communications means, the complete grid coordinates of the affected station are Item 6 of the MIJI 1 report.

The MIJI 2 Report. This is a complete report containing all details of the MIJI incident. Due to the number of items which require encryption when the report is transmitted over a nonsecure circuit, it is recommended that the report be delivered by messenger whenever possible. The higher headquarters' operations officer, intelligence officer, or the electronic warfare officer is responsible for ensuring that a complete message report of the incident is submitted to the Joint Command and Control Warfare Center (JC2WC) within 24 hours of the incident.

ELECTRONIC PROTECTION TECHNIQUES

Communications security (COMSEC) is an integral part of electronic protection. COMSEC is the protection resulting from all measures designed to deny unauthorized persons information of value that might be derived from the possession and study of telecommunications or to mislead unauthorized persons in their interpretation of the results of such possession and study. COMSEC includes transmission, cryptographic, emission, and physical security.

The goal of COMSEC is to protect friendly communications from enemy exploitation while ensuring unimpeded use of the electromagnetic spectrum. The organization must be able to employ communications equipment effectively in the face of enemy efforts.

COMSEC requirements must be integrated into communications systems planning and must focus on providing secure communications without impairing reliability or responsiveness. Modern communications equipment includes features such as an integrated encryption capability and frequency hopping capability, which contribute to communications protection. However, the security of our communications depends on the proper operation of communications equipment and adherence to proper procedures.

Transmission security

Transmission security (TRANSEC) is that component of COMSEC that results from all measures designed to protect transmissions from interception and exploitation by means other than cryptoanalysis. A message transmitted in the clear is the enemy's greatest source of information. After the enemy has intercepted your radio transmission, the enemy's language specialists will extract all possible intelligence from it. The enemy hopes to learn essential elements of friendly information (EEFI). Critical information that must be protected can be remembered by the key words SELDOM UP.

Each letter indicates a class of information as follows:

- Strength—number of personnel, size of unit.
- Equipment—type, quantity, condition.
- Logistics—procedure for resupply, depots.
- Disposition—were, what positions, map coordinates.
- Organization—how, what, chain of command, forces structure.
- Movement and morale—where, how, when and good or bad.
- Units—type, designation.
- Personalities—who, where.

Using TRANSEC is absolutely essential for the radio operator. When the radio must be used, keep transmission time to an absolute minimum (20 seconds absolute maximum: 15 seconds maximum preferred); preplan your messages to avoid compromising any essential element of information. If you must send EEFI items, use brevity lists, if possible, and also encrypt the message. These measures decrease your transmission, help prevent RDF, and deny the enemy valuable information. Included under transmission security are the authentication procedures that must be followed to protect against the enemy's IED. Every radio operator must be aware of the dangers of and guard against IED.

Strict radio discipline and adherence to authorized procedures are key to ensuring TRANSEC over SCR networks. SINCGARS radios should be operated in a frequency hopping mode to provide maximum protection against enemy EW capabilities. Other TRANSEC measures include—

- Well-trained operators thoroughly familiar with proper communications procedures and equipment operation. (This includes all Marines who may operate SCR, not just CIS personnel.)
- Avoidance of unauthorized transmission and testing and maximum use of data networks to minimize transmission time and opportunity for enemy direction finding.
- Use of transmitter, antenna, and power combinations that produce minimum wave propagation and emission intensity consistent with reliable communications.

- Strict adherence to authorized frequencies.
- Use of authentication systems to protect against imitative deception on nonsecure nets.
- Use of changing call signs and frequencies on nonsecure nets.
- Prompt response to and reporting of enemy jamming. (Operators should continue to operate on assigned frequencies in a secure mode, unless otherwise directed by a competent authority, and should attempt to work through the interference.)
- Strict adherence to all emission control (EMCON) restrictions and observance of radio silence.
- Use of communications means that do not radiate in the electromagnetic spectrum such as messengers, visual and sound signaling, and local wire loops.
- Use of terrain masking to shield transmission systems from enemy EW systems.
- Remoting of transmitters and avoiding the clustering of antennas.

Cryptosecurity

Cryptosecurity deals with codes, key lists, and communications security devices. This is the third line of defense for the radio operator. If the radio operator uses a security device on the radio, the enemy will not get anything for the language specialists to work on. However, do not get a false sense of security. The need for emission control and transmission security still exists—probably more so—because, if the enemy can't get information, the enemy might attempt to destroy or jam your station. Also, it is very important for all radio operators to use only authorized codes and to realize that using homemade codes is dangerous. Homemade codes offer no protection at all. Their use is not authorized and is a serious violation of security.

This also includes trying to "talk around" a classified or sensitive piece of information. The enemy intelligence personnel are not fools, and trying something like "talking around" critical information does more harm than good. If critical information must be transmitted, it should be encrypted or sent by secure means. In a situation when it is not possible to send by a secure means or to encrypt a message that must be sent, the

possibility of what friendly forces will lose against what the enemy could gain must be weighed. Other factors, such as how fast the enemy could react to the information and what delaying the message for encryption could mean, must also be considered.

Emission Security

Emission security (TEMPEST) is the component of COMSEC that results from all measures taken to deny unauthorized persons information of value that might be derived from interception and analysis of compromising emanations from cryptosecurity equipment and telecommunications systems. The operation of communications and information systems may result in unintentional electromagnetic emissions. Although tactical equipment is designed to reduce the possibility of such emissions, COTS equipment is not. Unintentional emissions are extremely susceptible to interception and analysis and may disclose classified information. Commanders must follow applicable regulations providing guidance on control and suppression of such emissions.

Physical Security

Physical security is the COMSEC component that results from all physical measures necessary to safeguard classified equipment, material, and documents from access or observation by unauthorized persons. The access to classified cryptographic information must be tightly controlled. When a commander or designated representative has determined that an individual has a need to know and is eligible for access, then access to classified cryptographic information will be formally authorized. The authorization process must include an introduction to the unique nature of cryptographic information, its unusual sensitivity, the special security regulations governing its handling and protection, and the penalties prescribed for its disclosure. Reportable violations include—

- Loss of material.
- Unauthorized viewing.
- Capture of individuals having access to COMSEC information.

Currently fielded COMSEC equipment is unclassified for external viewing when appropriate covers are in place and no keying material is visible. Consequently, the exposure of such equipment to casual viewing by uncleared personnel, whether by accident or as the result of operational necessity, does not constitute a reportable violation.

EP techniques are divided into two categories: preventive and remedial. Preventive EP are those procedures that can be used to avoid enemy EA attempts. Remedial EP apply to jamming only; there are no remedial measures once a unit has been intercepted, detected, or deceived.

ELECTRONIC WARFARE SUPPORT TECHNIQUES

Interception

The enemy is focused on intercepting your radio signal. To do this, all the enemy needs is a radio receiver that operates in the same mode and on the same frequency you are using to transmit. The mere fact that you are operating gives the enemy valuable information. It tells the enemy that you are in the area. By the number of stations operating on the same frequency, the enemy can estimate the size of the unit. If your net is operating in the clear, the enemy's language specialists can understand exactly what is said for even more information. When analyzing the traffic pattern, the enemy can figure out which station is the net control station (NCS) and identify the headquarters. Usually, in U.S. forces, the NCS is the radio used by the operations officer or section of the highest headquarters operating in the net. By further traffic analysis, the enemy can determine changes in the level of activity that could mean a movement or upcoming operation.

Radio Direction Finding

Interception is only one of the many dangers that the radio operator will face. After knowing friendly forces are in the area, the enemy will try to locate their position by using radio direction finding (RDF). A radio direction finder consists of a radio receiver, a directional antenna, and other specialized equipment. With RDF equipment, the approximate azimuth (i.e., bearing) to a transmitting radio can be determined. One

azimuth gives a general indication of direction. The intersection of two azimuths by different RDF stations is called a cut and gives a general indication of distance. The intersection of three or more azimuths is called a fix and gives a general location. The ideal fix is the exact intersection of three or more bearings. However, exact intersection is seldom achieved.

Terrain, weather, variations in radio wave propagation characteristics, the inherent RDF equipment, and operator inaccuracies, prevent an ideal fix. The fix that is obtained is called an actual fix. Although the actual fix may not be usable for immediate targeting purposes, it is more than enough for intelligence analysts to develop targeting data. Airborne direction finding is more accurate than ground-based direction finding but normally requires further analysis for targeting.

RDF ability to intercept electronics equipment emissions and determine a bearing depends on the power output of the targeting transmitter and its antenna radiation patterns. Experience indicates RDF accuracy of 500-meter (547-yard [yd]) circular error of probability (CEP) is considered a very good RDF fix. Normally, 50 percent of the CEPs are approximately 1,500 meters (1,640 yds) when the direction finder is located within 20 to 25 kilometers (12.4 to 15.5 mi) of the forward line of own troops (FLOT). Many threat forces will fire on a 1,500-meter (1,640-yd) CEP if they have sufficient massed artillery, and further analysis of terrain and radio intercept can reduce the target area or identify an important target.

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Appendix A

Map Coordinates

Field coordinates are often expressed in universal transverse mercator grid coordinates and usually consist of 6-digit numbers. The typical map used by the radio operator is a 1 to 50,000 scale topographic map that has grid lines drawn on it, which are 1,000 meters (1 kilometer or klick) apart. (See fig. A-1.)

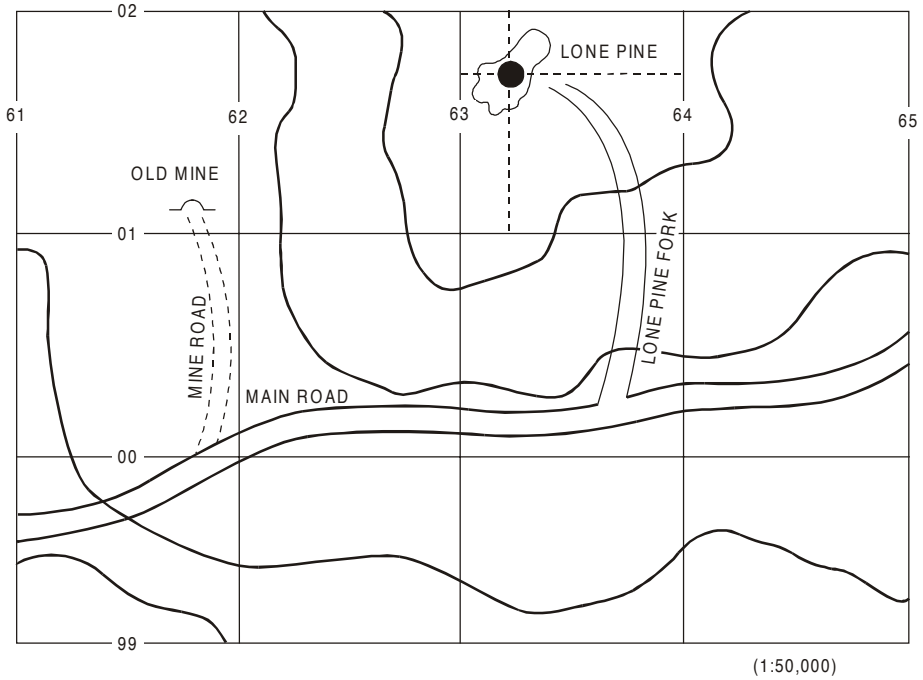

Figure A-1. Topographic Map.

To locate grid coordinate 632018, locate vertical line 63 (632018) in the figure and draw a vertical line 2/10 of the way (632018) between lines 63 and 64. Likewise, find horizontal line 01 (632018) and draw a

horizontal line 8/10 of the way (632018) between it and line 02. The two drawn lines intersect at Lone Pine.

Similarly, the intersection of the Lone Pine Fork with the Main Road in the figure would be represented as 638004, and the Old Mine would be 618012.

Although the use of the 6-digit number is generally sufficient for field use, a more exact coordinate will identify the 100,000 meter square and the grid zone designation to avoid confusion between different areas with identical grid line numbers.

A more detailed explanation of the complete coordinate may be found centered at the bottom of a 1 to 50,000 scale topographic map. The order in which the grids are read can be remembered by using the expression read-right-up.

Appendix B

Time Zones

The world is divided into 24 time zones, each one bearing a unique phonetic letter name (ROMEO, UNIFORM, etc.) or time zone number that must be applied to local time to arrive at the world standard time which is Coordinated Universal Time (UTC). This standard time is referred to in the Marine Corps as ZULU time. (See fig. B-1.) The time zones are roughly 15° apart in longitude.

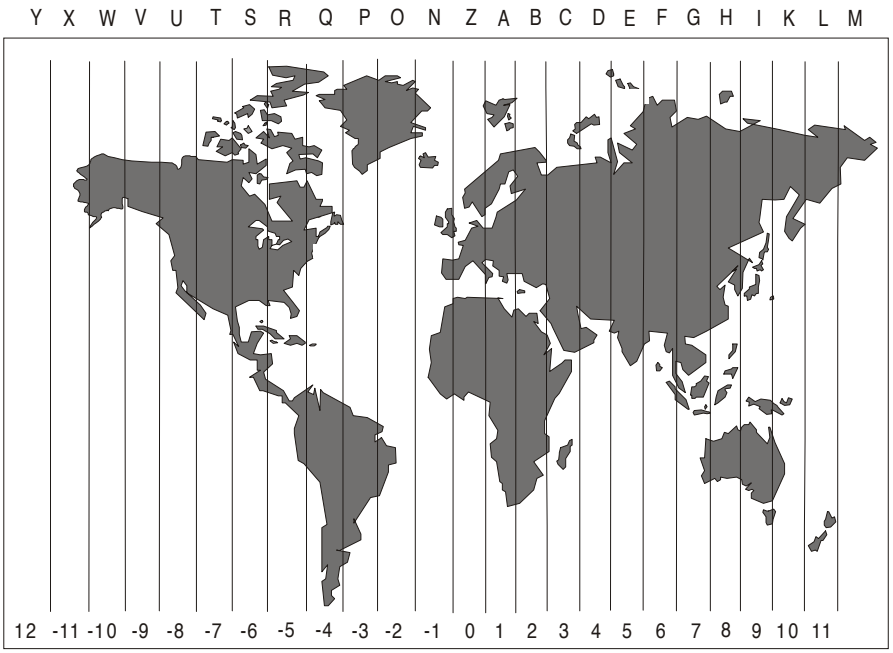

Figure B-1. Standard Time Zones of the World.

ZULU TIME = LOCAL TIME - TIME ZONE NUMBER.

LOCAL TIME = ZULU TIME + TIME ZONE NUMBER.

If a Marine is in the UNIFORM time zone, (also referred to as the PLUS 8 zone), 8 hours must be added to the local time to get ZULU time. (See table B-1.) On the other hand, in the ROMEO or PLUS 5 time zone, 1300 ZULU would equate to 0800 local. During the summer, however, if a Marine is located in an area where daylight savings time is observed, one hour must be subtracted from the time zone number; i.e., the PLUS 8 zone becomes the PLUS 7 zone for local time purposes.

Table B-1. CONUS Time Zones.

Civilian Time Zone	Military Time Zone	ZULU	Local Time	ZULU Time
EDT	QUEBEC	+4	0800 Q	1200 Z
EST CDT	ROMEO	+5	0700 R	1200 Z
CST MDT	SIERRA	+6	0600 S	1200 Z
MST PDT	TANGO	+7	0500 T	1200 Z
PST	UNIFORM	+8	0400 U	1200 Z

AUTODIN communications (i.e., worldwide) ZULU time should be used in all messages. Within the operational area, however, local time is usually used. To avoid confusion, the time zone should always be stated e.g., 1100 LOCAL or 1900 ZULU.

Appendix C

Prowords

Word or Phrase	Meaning
ALL AFTER	I refer to the portion of the message that follows.
ALL BEFORE	I refer to the portion of the message that precedes.
BREAK	I hereby indicate the separation of the text from other portions of the message. Or: I have completed the text of the message, signature follows, etc. (When break-in is permitted, receiving operator may interrupt the transmitting operator to request retransmission of a portion of a message. This proword is the interruption sign.)
CORRECTION	An error has been made in this transmission (or message indicated). The correct version is_____. That which follows is a corrected version in answer to your request for verification.
DISREGARD THIS TRANSMISSION	This transmission is in error. Disregard it. (This proword shall not be used to cancel any message that has been completely transmitted and for which receipt or acknowledgment has been received.)
DO NOT ANSWER	Stations called are not to answer this call receipt for this message, or otherwise to transmit in connection with this transmission. When this proword is employed, the transmission shall be ended with the proword *OUT*.

Word or Phrase	Meaning
EXECUTE	Carry out the purpose of the message or signal to which this applies. To be used only with the executive method.
EXECUTE TO FOLOW	Action on the message or signal which follows is to be carried out upon receipt of the proword *EXECUTE*. To be used only with the executive method.
EXEMPT	The addressee designations immediately following are exempted from the collective net call.
FIGURES	Numerals or numbers follow.
FLASH	Precedence *FLASH*.
FROM	The originator of this message is indicated by the address designation immediately following.
IMMEDIATE	Precedence *IMMEDIATE*.
INFO	The addressee designations immediately following are addressed for information.
I READBACK	The following is my response to your instructions to read back.
I SAY AGAIN	I am repeating transmission (or portion) indicated.
I SPELL	I shall spell the next word phonetically.
I VERIFY	The following message (or portion) has been verified at your request and is repeated. To be used only as a reply to verify.
MESSAGE FOLLOWS	A message which requires recording is about to follow. (Transmitted immediately after the call.)

Word or Phrase	Meaning
NUMBER	Station serial number.
OUT	This is the end of my transmission to you. No response is necessary.
OVER	This is the end of my transmission to you and a response is necessary. Go ahead and transmit.
PRIORITY	Precedence *PRIORITY.*
READ BACK	Repeat this entire transmission back to me exactly as received.
RELAY TO	Transmit this message to all addressees or to the address designations immediately following.
ROGER	I have received your last transmission satisfactorily.
ROUTINE	Precedence *ROUTINE*.
SAY AGAIN	Repeat all of your last transmission. Followed by identification data means: "Say again (portion indicated)." ("Repeat" is not used because it is the signal for naval gunfire and artillery to fire.)
SIGNALS FOLLOW	The groups which follow are taken from signal book. (This proword need not be used on nets primarily employed for conveying signals. It is intended for use when tactical signals are passed on nontactical nets.)
SILENCE, SILENCE, SILENCE	Cease transmission immediately. (Silence will be maintained until instructed to resume. When an authentication system is in force, transmissions imposing silence are to be authenticated.)

Word or Phrase	Meaning
SILENCE LIFTED	Resume normal transmissions. (Silence can be lifted only by the station imposing it or by higher authority. When an authentication system is in force, transmissions lifting silence are to be authenticated.)
SPEAK SLOWER	Your transmission is too fast. Reduce speed of transmission.
THAT IS CORRECT	You are correct, or what you have transmitted is correct.
THIS IS	This transmission is from the station whose designation immediately follows.
TIME	That which immediately follows is the time or date-time group of the message.
TO	The addressees whose designations immediately follow are to take action on this message.
UNKNOWN STATION	The identity of the station with whom I am attempting to establish communications is unknown.
VERIFY	Verify entire message (or portion indicated) with the originator and send correct version. To be used only at the discretion of or by the addressee to which the questioned message was directed.
WAIT	I must pause for a few seconds.
WAIT OUT	I must pause longer than a few seconds.

Word or Phrase	Meaning
WILCO	I have received your message, understand it, and will comply. (To be used only by the addressee. Since the meaning of ROGER is included in that of WILCO, the two prowords are never used together.)
WORD AFTER	I refer to the word that follows.
WORD BEFORE	I refer to the word that precedes.
WORDS TWICE	Communication is difficult. Transmit(ting) each phrase (or each code group) twice. This proword may be used as an order, request, or as information.
WRONG	Your last transmission was incorrect. The correct version is.

(reverse blank)

Appendix D

Phonetic Alphabet

Letter	Word	Pronunciation
A	ALFA	AL FAH
B	BRAVO	BRAH VOH
C	CHARLIE	CHAR LEE
D	DELTA	DELL TAH
E	ECHO	ECK OH
F	FOXTROT	FOKS TROT
G	GOLF	GOLF
H	HOTEL	HOH TELL
I	INDIA	IN DEE AH
J	JULIETT	JEW LEE ETT
K	KILO	KEY LOH
L	LIMA	LEE MAH
M	MIKE	MIKE
N	NOVEMBER	NOVEMBER
O	OSCAR	OSS CAH
P	PAPA	PAH PAH
Q	QUEBEC	KEH BECK
R	ROMEO	ROW ME OH
S	SIERRA	SEE AIR RAH
T	TANGO	TANG GO

Letter	Word	Pronunciation
U	UNIFORM	YOU NEE FORM
V	VICTOR	VIK TAH
W	WHISKEY	WISS KEY
X	XRAY	ECKS RAY
Y	YANKEE	YANG KEY
Z	ZULU	ZOO LOO

Appendix E

Phonetic Numerals

Number	Pronunciation
1	WUN
2	TOO
3	TREE
4	FOW-er
5	FIFE
6	SIX
7	SEV-en
8	AIT
9	NIN-er
0	ZERO

(reverse blank)

Appendix F

Prosigns

Prosign	Meaning
AA	All after.
AB	All before.
AR	End of transmission. No receipt required.
AS	I must pause for a few seconds.
AS AR	I must pause longer than a few seconds. Will call you back.
B	More to follow.
BT	Break.
C	Correct.
DE	This is.
EEEEEEEE	Error.
EEEEEEEE AR	This message is in error. Disregard it.
F	Do not answer.
FM	From.
G	Repeat this entire transmission back to me.
GR	Group count.
GRNC	The groups in this message have not been counted.
HM HM HM	Emergency silence sign-silence.

IMI ..... Repeat.

INFO ..... The address designations immediately following are addressed for information only.

INT ..... Interrogative.

IX ..... Action on the message or signal which follows is to be carried out upon receipt of execute.

J ..... Verify with originator and repeat.

K ..... This is the end of my transmission to you and a response is necessary.

NR ..... Numerals or numbers follow.

O ..... Immediate precedence.

P ..... Priority precedence.

R ..... Routine precedence.

T ..... Relay to.

TO ..... Action addressee.

WA ..... Word after.

WB ..... Word before.

XMT ..... Exempt.

Y ..... Emergency command precedence.

Z ..... Flash precedence.

Appendix G

Instructions for Preparing Field Messages

1. Place the protector insert under the message blanks to limit the number of copies produced. Retain one copy in the book as a file copy. Classify cover in accordance with contents (see fig. G-1).

2. Use BLOCK CAPITAL letters for all entries except the signature. Use necessary punctuation.

3. To assign precedence and classification use table on the next page.

4. If a classified message is to be transmitted, secure voice or in the clear, check appropriate block.

5. Show organization originating message in FM (from) block.

6. Show organization(s) for whom the message is intended in the TO block.

7. Block labeled date-time group is for communication personnel only.

8. Draft the message in brief but clear terms.

9. Message drafters are responsible for all message drafting functions to include the use of brevity codes.

PRECEDENCE TABLE

Z - FLASH
O- IMMEDIATE
P - PRIORITY
R - ROUTINE

SECURITY CLASSIFICATION TABLE

TOPSEC - TOP SECRET
SECRET - SECRET
CONF - CONFIDENTIAL
UNCLAS - UNCLASSIFIED

PROTECTOR INSERT

Place this under the last copy of each message written.

SAMPLE MESSAGE

PREC.	DTG	FM:	
TO:		INFO.	
BT	CLASS		
BT			
[X] SEND SECURE VOICE	[ ] SEND CLEAR	RELEASING OFFICER'S SIGNATURE	
TOR			TOD

Figure G-1. Sample Message.

Appendix H

Radio Log

Radio Circuit Log (Marine Corps)		
Circuit Bn TAC	Station Call A2C	Other stations J4Z M2P
Net ID/Frequency 556	Net Control Station X4L	
Operator PFC Jones, A	Net Call Sign S5F	
Supervisor Sgt. Smith, J	Page 01 of 10	Date 6 OCT 99

Time	Call	Transmission	End
0800Z	Assumed watch		
0801	B6D DE A2C A2C DE B30 A2C DE D5F	K K K	OVER
0804	B3D C3E D5F DE A2C	 AR	OUT
0808	A2C DE B3D B3D DE A2C A2C DE B3D	K K P 23140OZ SEP 99 C Files	OVER
0810	B3D DE A2C	AR	OUT
0812	B6D DE A2C	Abbreviated Calls Authorized AR	OUT
0815	C4E DE D5F	K	OVER
	DE E	K	
	DE F	MSG 0 231414Z SEP 99	
		FM D5F3	
		TO C4E3	
		BT	
		Classified (Use Actual Classification)	
		Report to CAR Dealer at 1000 Today for Liaison	
		BT	
		K	

(reverse blank)

Appendix I

Metric System Conversion Table

The basic unit of the metric system is the meter (m). The meter is 39.37 inches long. This is 3.37 inches longer than the English yard. Units that are multiples or fractional parts of the meter are designated by prefixes to the word "meter."

1 millimeter (mm) = 0.001 meter or 1/1000 meter
1 centimeter (cm)= 0.01 meter or 1/100 meter
1 decimeter (dm) = 0.1 meter of 1/10 meter
1 decameter (dkm) = 10 meters
1 hectometer (hm) = 100 meters
1 kilometer (km) = 1000 meters

The Metric Measurement in Most Common Use

10 millimeters = 1 centimeter
10 centimeters = 1 decimeter
10 decimeters = 1 meter
1000 meters = 1 kilometer

To Convert

Length	Multiply by
Inches to centimeters	2.54
Feet to meters	0.3048
Yards to meters	0.9144
Miles to kilometers	1.609
Millimeters to inches	0.03937
Centimeters to inches	0.3937
Decimeters to inches	3.937
Decimeters to feet	0.328
Meters to inches	39.37

Meters to feet	3.28
Meters to yards	1.0936
Decameters to feet	32.8
Hectometers to feet	328.1
Kilometers to feet	3281
Kilometers to yards	1093.6
Kilometers to miles	0.62

Examples:

To change 90 kilometers to miles: 90 x .62 = 55.8 miles
To change 90 kilometers to feet: 90 x 3281 = 295,290 feet
To change 50 yards to meters: 50 x 0.9144 = 45.72 meters

Appendix J

Authentication

Authentication systems are provided to prevent unauthorized enemy stations from entering friendly radio nets to disrupt or confuse operations. The only authentication systems authorized are those approved for use by the National Security Agency. If a special or emergency requirement arises, notify the CIS officer (G-6/S-6).

INSTRUCTIONS

There are two methods of authentication that are authorized for use: challenge and reply authentication and transmission authentication. The operational distinction is that challenge and reply requires two-way communications, whereas transmission authentication does not.

Challenge and Reply Authentication

Challenge and reply authentication will be used whenever possible. The called party will always make the first challenge. Besides validating the authenticity of the calling station, this practice prevents an enemy operator from entering a net to obtain correct authentication responses for use in another net. The party making the call may counterchallenge the called party using a different challenge.

Note: In challenge and reply authentication, only the station responding is verified. Do not accept a challenge as an authentication.

When a caller desires authentication, he must invite a challenge by stating that he is prepared to authenticate.

Another challenge should be made if an incorrect reply is received, if a standby is requested, or if an unusual delay occurs between challenge and reply.

Users will occasionally misauthenticate because of such problems as having the wrong system, misreading the table, etc. The challenging station should attempt to pinpoint the difficulty and then rechallenge.

Never give the challenge and reply in the same transmission (self authentication).

Transmission Authentication

Transmission authentication is used to validate the authenticity of the message when it is impossible or impractical to use challenge and reply authentication (see table J-1).

For Instructional Purposes Only

Table J-1. Sample AKAC-874 Transmission Authentication.

	18	19	20	21	22	23
0	SD	IU	PY	UT	JG	LJ
2	FG	KG	AG	RG	YR	HF
4	VF	DY	KU	JM	RW	FS
6	HD	ML	DT	SC	DA	SA
8	RD	NB	MG	GR	LJ	QW
10	BJ	FR	SR	EF	MB	TP
12	JH	SP	PO	QS	VX	KU
14	TY	OL	RM	OL	FS	DR
16	DR	IJ	AO	MJ	HD	RT
18	JH	TL	KY	BG	UY	NY
20	SD	WM	SR	DE	GC	TR
22	MJ	AP	GH	FD	JG	OU
24	BN	PC	FI	KI	RW	TM

WHEN TO AUTHENTICATE

Transmission should be authenticated—

- When any station suspects imitative deception on any circuit; e.g., when contacting a station following one or more unsuccessful attempts to contact that station.
- When any station is challenged or requested to authenticate. This is not to be interpreted as requiring stations to break an imposed silence for the sole purpose of authenticating.
- When directing radio silence, listening silence, or requiring a station to break an imposed silence.
- When transmitting contact and amplifying reports in plain language.
- When transmitting operating instructions that affect the military situation; e.g., closing down a station or watch, changing frequency other than normal scheduled changes, directing establishment of a special communication guard, requesting artillery fire support, directing relocation of units, etc.
- When transmitting a plain language cancellation.
- When making initial radio contact or resuming contact after prolonged interruptions.
- When transmitting to a station that is under radio listening silence.
- When authorized to transmit a classified message in the clear.
- When forced, because of no response by the called station to send a message in the blind (transmission authentication).

Note: Authentication is not required when making initial contact after a scheduled call sign and frequency change since only bona fide stations will know their assigned call sign and frequency for the time period in use.

For detailed authentication instructions refer to the automated communications-electronics operations instructions.

(reverse blank)

Appendix K

International Morse Code

Dots and dashes are used in various distinctive combinations to represent the letters of the alphabet, the numerals from 0 to 9, and the prosigns (see fig K-1). The dots and dashes of the Morse Code are produced by keying a transmitter and causing it to transmit short and long signals. The dash is 3 times the length of the dot. The dots and dashes used for a letter are spaced from each other by a period of time equal in length to one dot. Letters are spaced from each other by a period of time equal to three dots. Words are spaced by a period of time equal to seven dots.

a. Letters

A • —	J • — — —	S • • •
B — • • •	K — • —	T —
C — • — •	L • — • •	U • • —
D — • •	M — —	V • • • —
E •	N — •	W • — —
F • • — •	O — — —	X — • • —
G — — •	P • — — •	Y — • — —
H • • • •	Q — — • —	Z — — • •
I • •	R • — •	

b. Figures

1 • — — — —	5 • • • • •	9 — — — — •
2 • • — — —	6 — • • • •	0 — — — — —
3 • • • — —	7 — — • • •	
4 • • • • —	8 — — — • •	

Figure K-1. Morse Code.

(reverse blank)

Appendix L

Frequency Prediction Means

Radio signals propagate from a transmitter to a receiver in different ways depending on the selected frequency of the radio. The two available resources for Marines to predict the best frequency to propagate over a given path are to use the Joint Spectrum Center (JSC) or the Marine Corps' system planning, engineering, and evaluation device (SPEED). See figure L-1 on page L-2.

JSC is a Department of Defense agency that is responsible for supplying the electromagnetic analysis to the uniformed services. Located in Annapolis, Maryland, JSC can perform a variety of propagation predictions on all ranges of frequencies. Once a quarter, JSC publishes a list of HF frequency predictions for selected paths that the Marine Corps has requested. A compilation of frequency predictions are held in the G-6 offices of the major subordinate commands within the Marine Corps.

The second method is SPEED. SPEED is computer software that allows communications system planners to do rapid analysis of radio signal propagation. The SPEED comes with electronic maps of the Earth which allow true line of sight data for the planners, as well as HF propagation prediction. SPEED is available down to the battalion level.

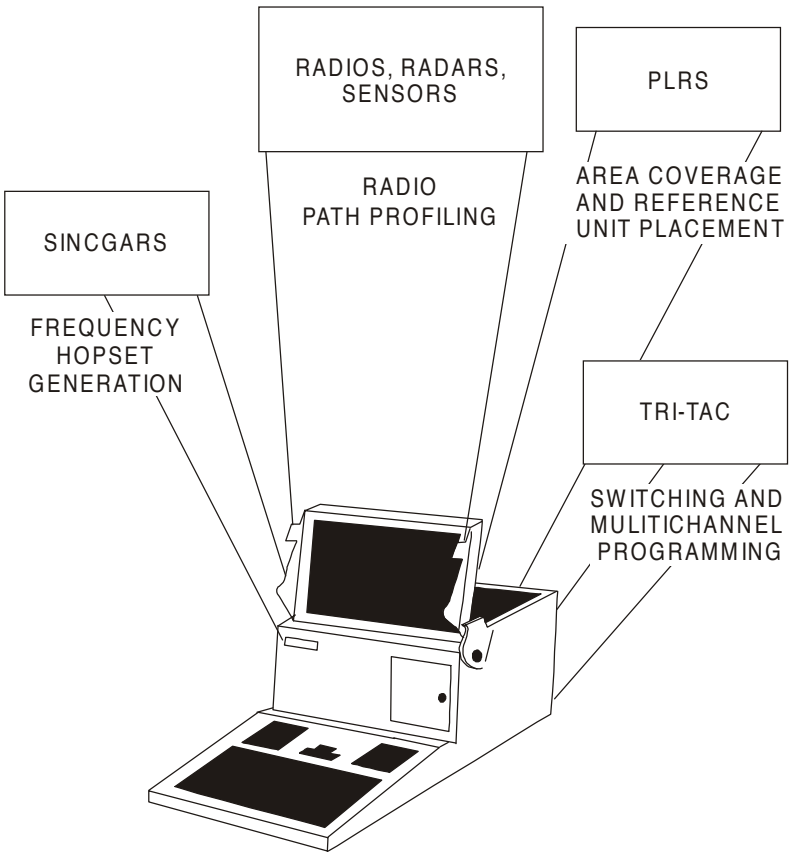

Figure L-1. SPEED.

Appendix M

Position and Navigation Systems

Global positioning system (GPS) and position location reporting system (PLRS) are new systems that radio operators may be required to operate.

GLOBAL POSITIONING SYSTEM

Global positioning system is a space-based navigation system designed to provide 24-hour continuous worldwide, all-weather precise position, and time measurement. The GPS consists of a space segment (satellite), control segment (monitors stations on Earth), and user segment (GPS receivers). The system operates by satellites sending out two signals on nonchanging frequencies. The GPS receiver receives the signals transmitted by the satellites and computes the users position. GPS (by being an all-weather, jam-resistant, continuous system) gives users highly accurate navigation; worldwide, three-dimensional position or location velocity; and time information. As a passive, receive-only system, GPS can be employed at the individual level in such nondescript terrain as jungles, mountain ranges, or deserts. (See fig. M-1 on page M-2.)

POSITION LOCATION REPORTING SYSTEM

PLRS is a system of UHF radios, signal and message processors, and user input and output devices configured as master stations and basic units that provide the user of the system with position navigation information and limited digital communications. PLRS works by the individual units in the network time sharing a single frequency band. As the master station records the arrival time of the signal bursts from each user unit at given locations, the range between sender and receiver can be computed. If units are beyond line of sight for the master station, then PLRS will automatically enable any user unit to serve as an automatic relay. The full PLRS performance can be provided over a 47- by 47-kilometer operating area. This operating area can be extended to a 300-by 300-kilometer area through the use of airborne relays provided by PLRS-equipped aircraft.

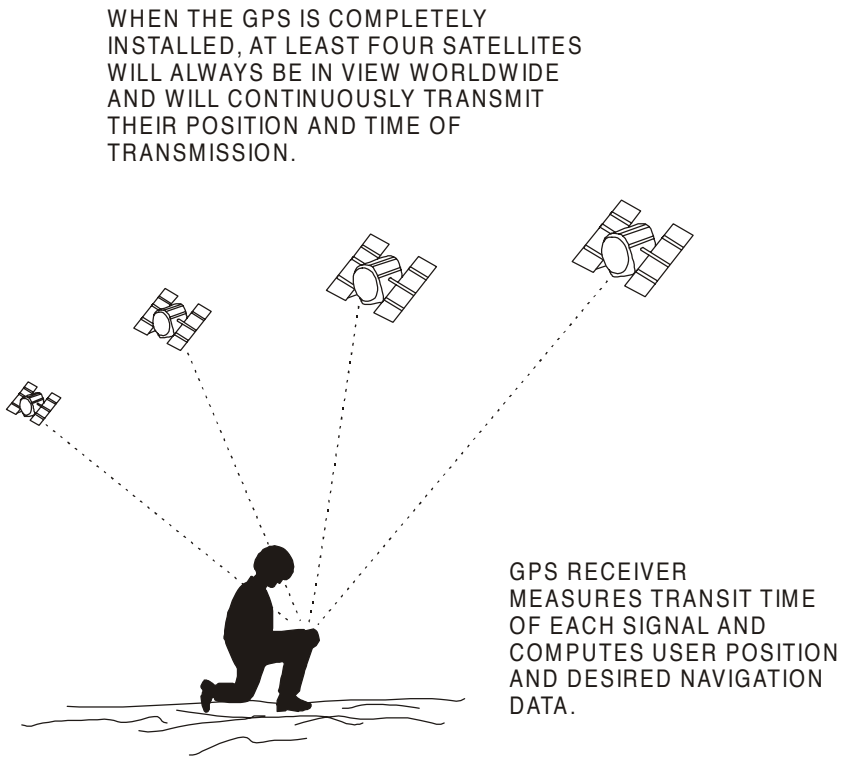

Figure M-1. Global Positioning System.

ENHANCED POSITION REPORTING SYSTEM

Enhanced position location reporting system (EPLRS) shares many characteristics with PLRS, but provides a significant increase in data communications capability over PLRS. Various data rates supporting a variety of broadcast and point-to-point modes are currently available. EPLRS will provide a dedicated data communications capability between regiment and battalion tactical data networks (TDNs) within the ground combat element, when fielded in FY-00. This network will also be extended to lower echelons throughout the MAGTF. EPLRS can also serve as a source for automated friendly position location information and navigation information in a hybrid community with PLRS, though data throughput is reduced.

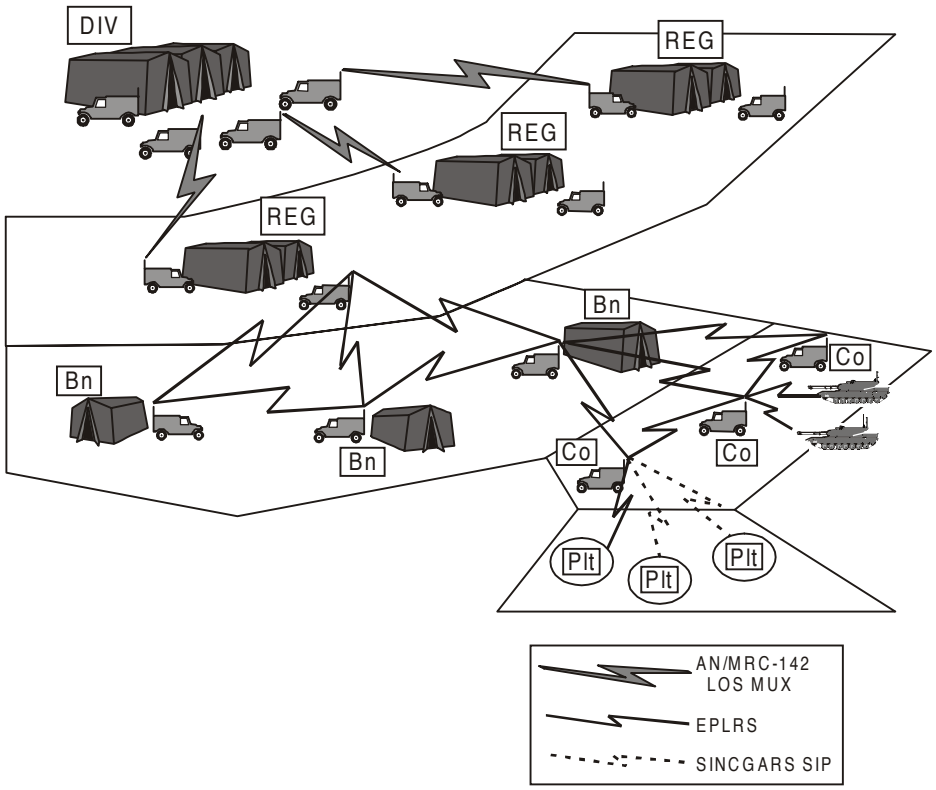

Figure M-2. EPLRS Concept of Employment.

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Appendix N

Size of Dipole and Inverted L Antennas

ACCORDING TO WAVELENGTH MEASUREMENTS

Size of dipole and inverted L antennas according to wavelength measurements (in feet and inches).

Table N-1. Size of Antennas.

Frequency	1/2 Wavelength	1/4 Wavelength	1/8 Wavelength	2 1/2 Wavelength (MHz)
2.00	225 ft 6 in	112 ft 9 in	56 ft 4 1/2 in	1,127 ft 6 in
2.50	180 ft 4 3/4 in	90 ft 2 1/3 in	45 ft 1 in	901 ft 11 3/4 in
3.00	150 ft 4 in	75 ft 2 in	37 ft 7 in	751 ft 8 in
3.50	128 ft 10 1/4 in	64 ft 5 in	32 ft 2 in	644 ft 3 3/8 in
4.00	112 ft 9 in	56 ft 4 1/2 in	28 ft 21/2 in	563 ft 9 in
4.50	100 ft 2 2/3 in	50 ft 11/3 in	25 ft 2/3 in	501 ft 1 1/3 in
5.00	90 ft 2 1/3 in	45 ft 1 in	22 ft 6 2/3 in	450 ft 11 3/4 in
5.50	82 ft	41 ft	20 ft 6 in	410 ft
6.00	75 ft 2 in	37 ft 7 in	18 ft 9 1/2 in	375 ft 10 in
6.50	69 ft 4 2/3 in	34 ft 8 1/3 in	17 ft 4 in	346 ft 11 in
7.00	64 ft 5 in	32 ft 2 1/2 in	16 ft 1 1/4 in	322 ft 1 3/4 in
7.50	60 ft 1 2/3 in	30 ft 1 in	15 ft 1/3 in	300 ft 8 in
8.00	56 ft 4 1/2 in	28 ft 2 1/4 in	14 ft 1 in	281 ft 10 1/2 in
8.50	53 ft 3/4 in	26 ft 6 1/8 in	13 ft 3 1/8 in	265 ft 3 1/2 in
9.00	50 ft 1 1/3 in	25 ft 2/3 in	12 ft 1/3 in	250 ft 6 2/3 in
9.50	47 ft 5 2/3 in	23 ft 8 3/4 in	11 ft 10 1/3 in	237 ft 4 3/8 in

Table N-1. Size of Antennas (continued).

Frequency	1/2 Wavelength	1/4 Wavelength	1/8 Wavelength	2 1/2 Wavelength (MHz)
10.00	45 ft 1 in	22 ft 6 2/3 in	11 ft 1/4 in	225 ft 6 in
10.50	42 ft 11 2/3 in	21 ft 5 3/4 in	10 ft 8 3/4 in	214 ft 9 1/8 in
11.00	41 ft	20 ft 6 in	10 ft 3 in	205 ft
11.50	39 ft 2 2/3 in	19 ft 7 1/2 in	9 ft 9 2/3 in	196 ft 1 in
12.00	37 ft 7 in	18 ft 9 1/2 in	9 ft 4 3/4 in	187 ft 11 in
12.50	36 ft 1 in	18 ft 1/2 in	9 ft 1/4 in	180 ft 4 3/4 in
13.00	34 ft 8 1/3 in	17 ft 4 in	8 ft 8 in	173 ft 5 1/2 in
13.50	33 ft 5 in	16 ft 8 1/2 in	8 ft 4 1/4 in	167 ft 3/8 in
14.00	32 ft 2 1/2 in	16 ft 1 1/4 in	8 ft 2/3 in	161 ft 3/4 in
14.50	31 ft 1 1/4 in	15 ft 6 2/3 in	7 ft 9 1/3 in	155 ft 6 1/4 in
15.00	30 ft 1 in	15 ft 1/3 in	7 ft 6 1/4 in	150 ft 4 in
15.50	29 ft 1 1/8 in	14 ft 6 1/2 in	7 ft 3 1/4 in	145 ft 5 7/8 in
16.00	28 ft 2 1/4 in	14 ft 1 1/8 in	7 ft 1/2 in	140 ft 11 1/4 in
16.50	27 ft 4 in	13 ft 8 in	6 ft 10 in	136 ft 8 in
17.00	26 ft 6 1/3 in	13 ft 3 1/8 in	6 ft 7 1/2 in	132 ft 7 3/4 in
17.50	25 ft 9 1/4 in	12 ft 10 2/3 in	6 ft 5 1/3 in	128 ft 10 1/4 in
18.00	25 ft 2/3 in	12 ft 6 1/3 in	6 ft 3 1/8 in	125 ft 3 1/2 in
18.50	24 ft 4 1/2 in	12 ft 2 1/4 in	6 ft 1 1/8 in	121 ft 10 3/4 in
19.00	23 ft 8 3/4 in	11 ft 10 3/8 in	5 ft 11 1/4 in	118 ft 8 1/8 in
19.50	23 ft 1 1/2 in	11 ft 6 3/4 in	5 ft 9 3/8 in	115 ft 7 2/3 in
20.00	22 ft 6 1/2 in	11 ft 3 1/4 in	5 ft 7 2/8 in	112 ft 9 in
20.50	22 ft	11 ft	5 ft 6 in	110 ft
21.00	21 ft 5 3/4 in	10 ft 8 1/8 in	5 ft 4 3/8 in	107 ft 4 1/2 in
21.50	20 ft 11 3/4 in	10 ft 5 1/8 in	5 ft 3 in	104 ft 10 2/3 in
22.00	20 ft 6 in	10 ft 3 in	5 ft 1 1/2 in	102 ft 6 in
22.50	20 ft 1/2 in	10 ft 1/4 in	5 ft 1/8 in	100 ft 2 2/3 in

Table N-1. Size of Antennas (continued).

Frequency	1/2 Wavelength	1/4 Wavelength	1/8 Wavelength	2 1/2 Wavelength (MHz)
23.00	19 ft 7 1/3 in	5 ft 9 2/3 in	4 ft 10 1/8 in	98 ft 1/2 in
23.50	19 ft 2 1/4 in	9 ft 7 1/8 in	4 ft 9 in	95 ft 11 1/2 in
24.00	18 ft 9 1/2 in	9 ft 4 3/4 in	4 ft 8 3/8in	93 ft 11 1/2 in
24.50	18 ft 5 in	9 ft 2 3/8 in	4 ft 7 1/4 in	92 ft 1/2 in
25.00	18 ft 1/2 in	9 ft 1/4 in	4 ft 6 1/8 in	90 ft 2 3/8 in
25.50	17 ft 8 1/4 in	8 ft 10 1/8 in	4 ft 5 in	88 ft 5 1/8 in
26.00	17 ft 4 1/8 in	8 ft 8 in	4 ft 4 in	85 ft 8 1/8 in
26.50	17 ft 1/4 in	8 ft 6 1/8 in	4 ft 3 in	85 ft 1 1/8 in
27.00	16 ft 8 3/8 in	8 ft 4 1/4 in	4 ft 2 1/8 in	83 ft 6 in
27.50	16 ft 4 3/4 in	8 ft 2 3/8 in	4 ft 1 1/8 in	81 ft 11 1/8 in
28.00	16 ft 1 1/4 in	8 ft 2/3 in	4 ft 1/3 in	80 ft 6 3/8 in
28.50	15 ft 9 1/8 in	7 ft 11 in	3 ft 11 1/2 in	79 ft 1 1/2 in
29.00	15 ft 6 2/3 in	7 ft 9 1/3 in	3 ft 10 2/3 in	77 ft 9 1/8 in
29.50	15 ft 3 1/2 in	7 ft 7 3/4 in	3 ft 10 in	76 ft 5 1/4 in
30.00	15 ft 3/8 in	7 ft 6 1/8 in	3 ft 9 in	75 ft 2 in

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Appendix O

Field Repair and Expedients

Section I. Antenna Repair

Antennas are sometimes broken or damaged, causing either a communications failure or poor communications. If a spare is available, replace the damaged antenna. When there is no spare, you may have to construct an emergency antenna. The following paragraphs are suggestions on repairing antennas and antenna supports and on constructing and adjusting emergency antennas.

REPAIR TECHNIQUES

Whip Antennas

When a whip antenna is broken into two sections, the portion of the antenna that is broken off can be connected to the portion attached to the base by joining the sections as shown in figure O-1 on page O-2. Use the method illustrated in figure O-1A, when both parts of the broken whip are available and usable. Use the method shown in figure O-1B when the portion of the whip that was broken off is lost, or when the whip is badly damaged. To restore the antenna to its original length, add a piece of wire that is nearly the same length as the missing part of the whip. Then, lash the pole support securely to both sections of the antenna. Clean the two antenna sections thoroughly to ensure good contact before connecting them to the pole support. If possible, solder the connections.

Wire Antennas

Emergency repair of a wire antenna may involve the repair or replacement of the wire used as the antenna or transmission line; or, the repair or replacement of the assembly used to support the antenna.

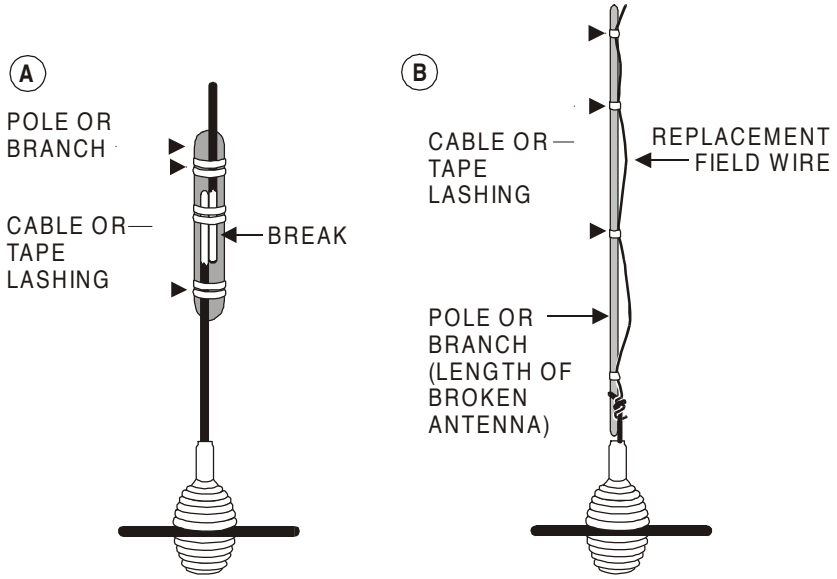

Figure O-1. Emergency Repair of Broken Whip.

When one or more wires of an antenna are broken, the antenna can be repaired by reconnecting the broken wires. To do this, lower the antenna to the ground, clean the ends of the wires, and twist the wires together. Whenever possible, solder the connection.

If the antenna is damaged beyond repair, construct a new one. Make sure that the lengths of the wires of the substitute antenna are the same length as the original.

Antenna supports may also require repair or replacement. A substitute item may be used in place of a damaged support and, if properly insulated, can be of any material of adequate strength. If the radiating element is not properly insulated, field antennas may be shorted to ground and rendered ineffective. Many commonly found items can be used as field expedient insulators. The best of these items are plastic or glass, to include plastic spoons, buttons, bottle necks, and plastic bags. Less effective than plastic or glass but better than no insulators at all are wood and rope, or both, in that order. The radiating element—the actual antenna wire—should touch only the antenna terminal and be physically

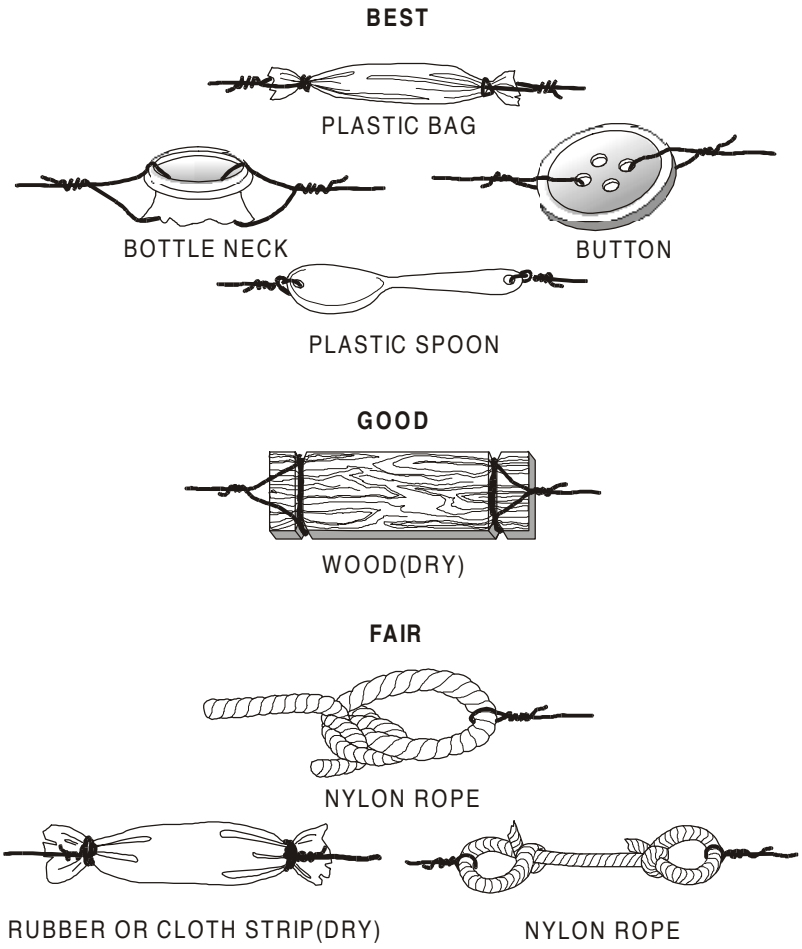

Figure O-2. Improvised Insulators.

separated from all other objects, other than the supporting insulator. Figure O-2 shows various emergency insulators.

Guys

Lines used to stabilize the supports for an antenna are called guys. These lines are usually made of wire, manila rope, or nylon rope. If a rope breaks, it may be repaired by tying the two broken ends together. If the rope is too short after the tie is made, it can be lengthened by adding another piece of dry wood or cloth. If a guy wire breaks, it can be

replaced with another piece of wire. Figure O-3 shows a method of repairing a guy line with a spoon.

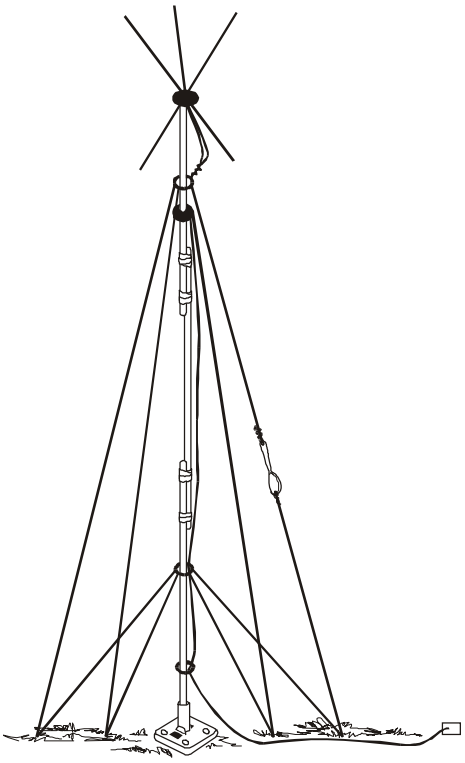

Figure O-3. Repaired Guy Lines and Masts.

Masts

Some antennas are supported by masts. If a mast breaks, it can be replaced with one of same length. If long poles are not available as replacements, short poles may be overlapped and lashed together with rope or wire to provide a pole of the required length. Figure O-3 shows a method of making an emergency repair to masts.

TIPS FOR CONSTRUCTION AND ADJUSTMENT

Constructing the Antenna

The best kinds of wire for antennas are copper and aluminum. In an emergency, however, use any type that is available.

The length of most antennas is critical. The emergency antenna should be the same length as the antenna it replaces.

Antennas supported by trees can usually survive heavy wind storms if the trunk of a tree or a strong branch is used as a support. To keep the antenna taut and to prevent it from breaking or stretching as the trees sway, attach a spring or old inner tube to one end of the antenna. Another technique is to pass a rope through a pulley or eyehook, attach the rope to the end of the antenna, and load the rope with a heavy weight to keep the antenna tightly drawn.

Guys used to hold antenna supports are made of rope or wire. To ensure that the guys made of wire will not affect the operation of the antenna, cut the wire into several short lengths and connect the pieces with insulators.

Adjusting the Antenna

An improvised antenna may change the performance of a radio set. Use the following methods to determine if the antenna is operating properly.

A distant station may be used to test the antenna. If the signal received from this station is strong, the antenna is operating satisfactorily. If the signal is weak, adjust the height and length of the antenna and the transmission line to receive the strongest signal at a given setting on the volume control of the receiver. This is the best method of tuning an antenna when transmission is dangerous or forbidden.

In some radio sets, the transmitter is used to adjust the antenna. First, set the controls of the transmitter in the proper position for normal operation; then, tune the system by adjusting the antenna height, the antenna

length, and the transmission line length to obtain the best transmission output.

Impedance-matching a load to its source is an important consideration in transmissions' systems. If the load and source are mismatched, part of the power is reflected back along the transmission line towards the source. This reflection not only prevents maximum power transfer, but can also be responsible for erroneous measurements of other parameters, or even cause circuit damage in high-power applications.

The power reflected from the load interferes with the incident (i.e., forward) power, causing standing waves of voltages and current to exist along the line. The ratio of standing-wave maxima to minima is directly related to the impedance mismatch of the load; therefore the standing-wave ratio (SWR) provides the means of determining impedance and mismatch.

WARNING

SERIOUS INJURY OR DEATH CAN RESULT FROM CONTACT WITH THE RADIATING ANTENNA OF A MEDIUM- OR HIGH-POWER TRANSMITTER. TURN THE TRANSMITTER OFF WHILE MAKING ADJUSTMENTS TO THE ANTENNA.

Section II. Field Expedient Antennas

SINCGARS VHF radios provide the primary means of communications for Marine Corps forces around the world. The SINCGARS radio operates in both single-channel and frequency hopping modes. It is important for CIS personnel to remember that when using the SINCGARS radio in the frequency hopping mode, field expedient VHF antennas should not be used. CIS personnel should only use the whip antenna or the OE-254 antenna when operating in the frequency hopping mode.

HIGH FREQUENCY, FIELD EXPEDIENT OMNIDIRECTIONAL ANTENNAS

Vertical antennas are omnidirectional; i.e., they transmit and receive equally well in all directions. Most manpack portable radios use a vertical whip antenna. A vertical antenna can be improvised by using a metal pipe or rod of the correct length, held erect by means of guys. The lower end of the antenna should be insulated from the ground by placing it on insulating material. A vertical antenna may also be a wire, supported by a tree or a wooden pole (see fig. O-4). For short, vertical

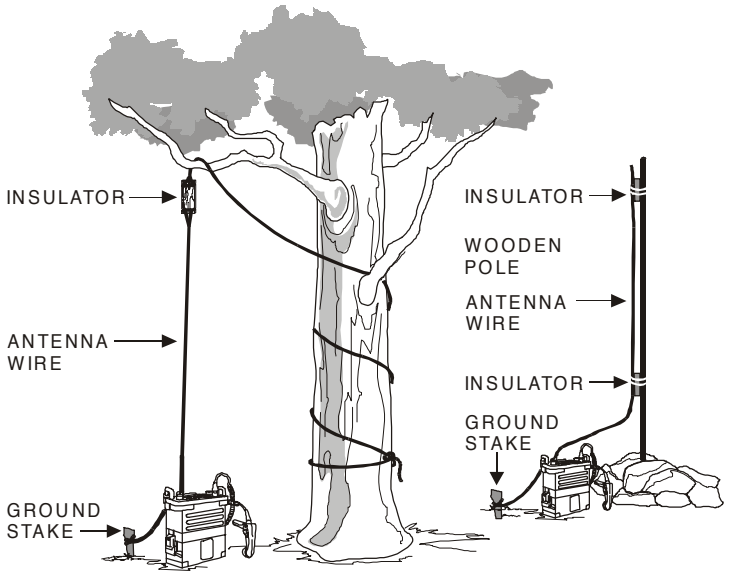

Figure O-4. Field Substitutes for Support of Vertical Wire Antennas.

antennas, the pole may be used without guys (if properly supported at the base). If the length of the vertical mast is not long enough to support the wire upright, it may be necessary to modify the connection at the top of the antenna.

End-Fed Half-Wave Antenna

An emergency, end-fed half-wave antenna can be constructed from available materials such as field wire, rope, and wooden insulators. The electrical length of this antenna is measured from the antenna terminal on the radio set to the far end of the antenna (see fig. O-5). Construct the antenna longer than necessary, then shorten it, as required, until best results are obtained. The ground terminal of the radio set should be connected to a good Earth ground for this antenna to function efficiently.

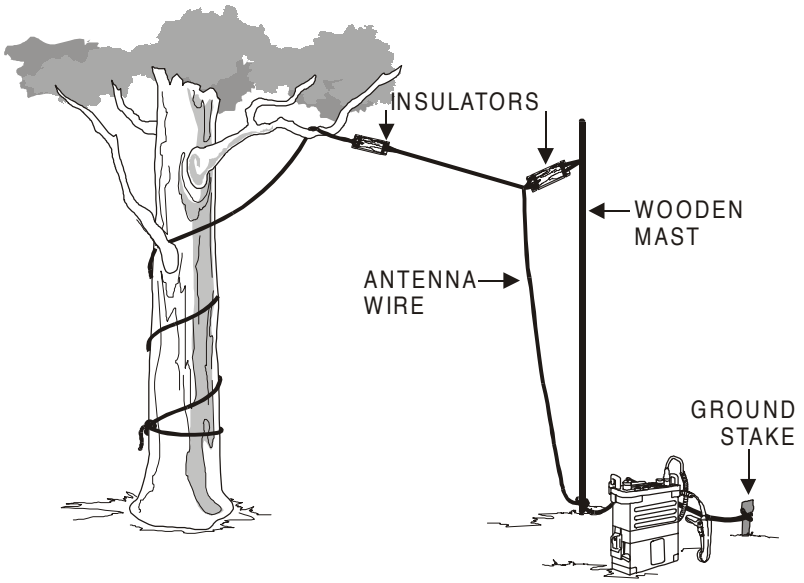

Figure O-5. End-Fed Half-Wave Antenna.

Center-Fed Doublet Antenna

The center-fed doublet is a half-wave antenna consisting of two, quarter-wavelength sections on each side of the center. Construction of an improvised doublet antenna for use with FM radios is shown in figure O-6.

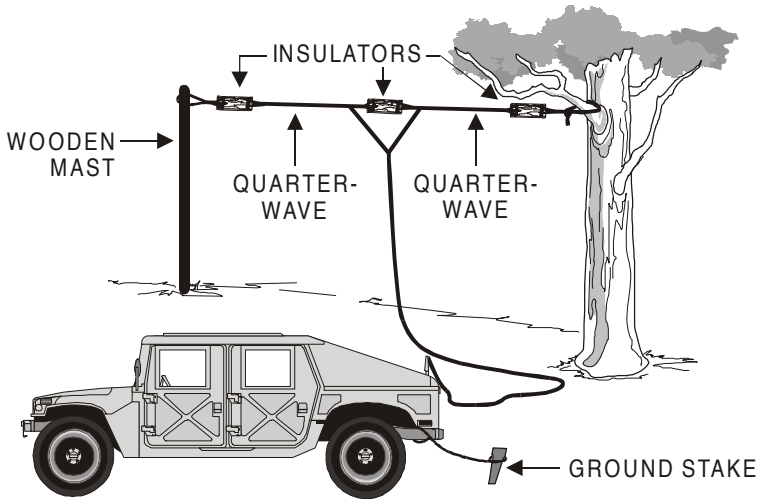

Figure O-6. Half-Wave Doublet Antenna.

Doublet antennas are directional broadside to their length, which makes the vertical doublet antenna essentially omnidirectional. This is because the radiation pattern is doughnut shaped. The horizontal doublet antenna is bidirectional.

The length of a half-wave antenna may be computed by using the formula in Chapter 4. Cut the wires as closely as possible to the correct length because the length of the antenna wires is important.

A transmission line is used for conducting electrical energy from one point to another, and it is used to transfer the output of a transmitter to an antenna. Although it is possible to connect an antenna directly to a transmitter, the antenna generally is located some distance away. In a vehicular installation, for example, the antenna is mounted outside, and the transmitter is inside the vehicle. A transmission line, therefore, is necessary as a connecting link.

Center-fed half-wave FM antennas can be supported entirely by pieces of wood. A horizontal antenna of this type is shown in figure O-7A; a vertical antenna in figure O-7B on page O-10. These antennas can be rotated to any position to obtain the best performance. If the antenna is erected vertically, the transmission line should be brought out horizontally from the antenna for a distance equal to at least one-half of the antenna's length before it is dropped down to the radio set.

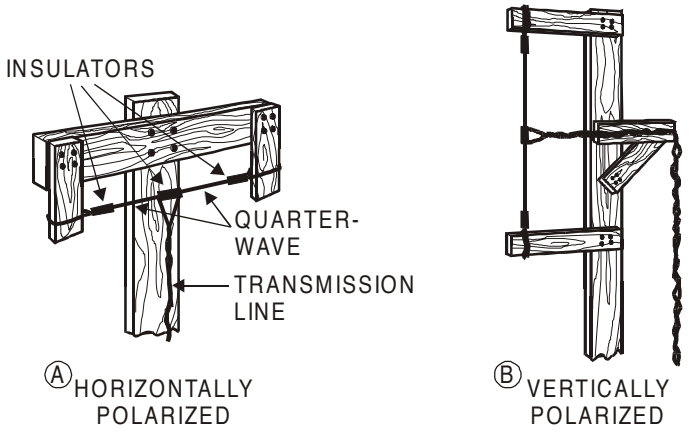

Figure O-7. Center-Fed Half-Wave Antenna.

A similar arrangement for a short, center-fed half-wave antenna is shown in figure O-8. The ends of this antenna are connected to a piece of dry wood, such as a bamboo pole, and the bend in the pole holds the antenna wire straight. Another pole, or bundle of poles, serves as the mast.

Figure O-9 shows an improvised vertical half-wave antenna. This technique is used primarily with FM radios. It is effective in heavily wooded

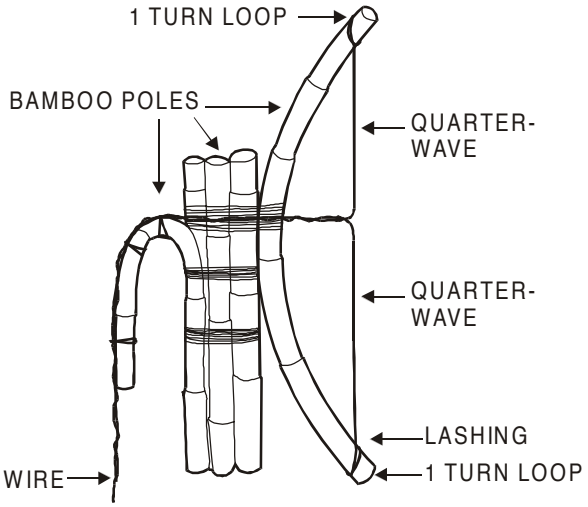

Figure O-8. Bent Bamboo Antenna.

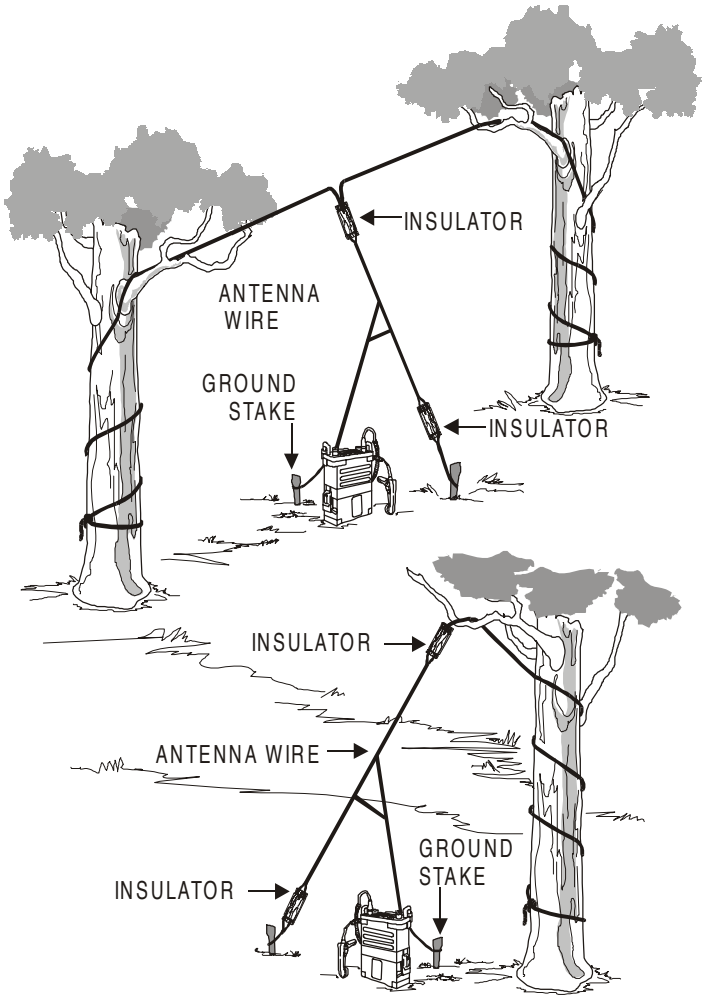

Figure O-9. Improvised Vertical Half-Wave Antennas.

areas to increase the range of portable radios. The top guy wire can be connected to a limb or passed over the limb and connected to the tree trunk or a stake.

HIGH FREQUENCY, FIELD EXPEDIENT DIRECTIONAL ANTENNAS

The vertical half-rhombic antenna (fig. O-10 on page O-12) and the long-wire antenna (fig. O-11 on page O-12) are two field expedient,

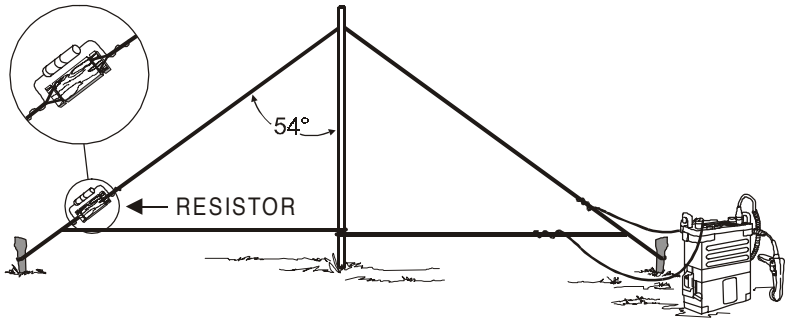

Figure O-10. Vertical Half-Rhombic Antenna.

directional antennas. These antennas consist of a single wire, preferably two or more wavelengths long, supported on poles at a height of 3 to 7 meters (10 to 20 feet) above the ground. The antennas will, however, operate satisfactorily as low as 1 meter (approximately 3 feet) above the ground. The far end of the wire is connected to ground through a noninductive resistor of 500 to 600 ohms. Use a resistor rated at least one-half the wattage output of the transmitter to ensure the resistor is not burned out by the output power of the transmitter. A reasonably good ground, such as a number of ground rods or a counterpoise, should be used at both ends of the antenna. The radiation pattern is directional. The antennas are used primarily for either transmitting or receiving high frequency signals.

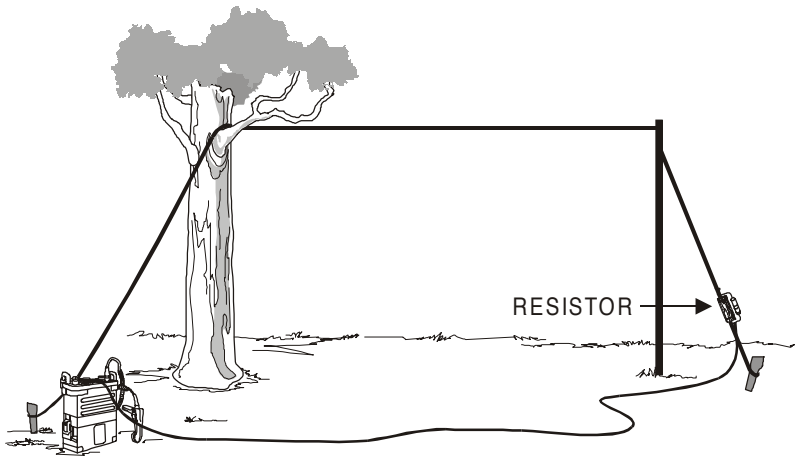

Figure O-11. Long-Wire Antenna.

The Vee antenna is another field expedient, directional antenna. It consists of two wires forming a Vee with the open area of the Vee pointing toward the desired direction of transmission or reception (see fig. O-12). To make construction easier, the legs may slope downward from the apex of the Vee (this is called a sloping Vee antenna [see fig. O-13 on page O-14]).

The angle between the legs varies with the length of the legs in order to achieve maximum performance.

When the antenna is used with more than one frequency or wavelength, use an apex angle that is midway between the extreme angles determined by the chart.

To make the antenna radiate in only one direction, add noninductive terminating resistors from the end of each leg (not at the apex) to ground.

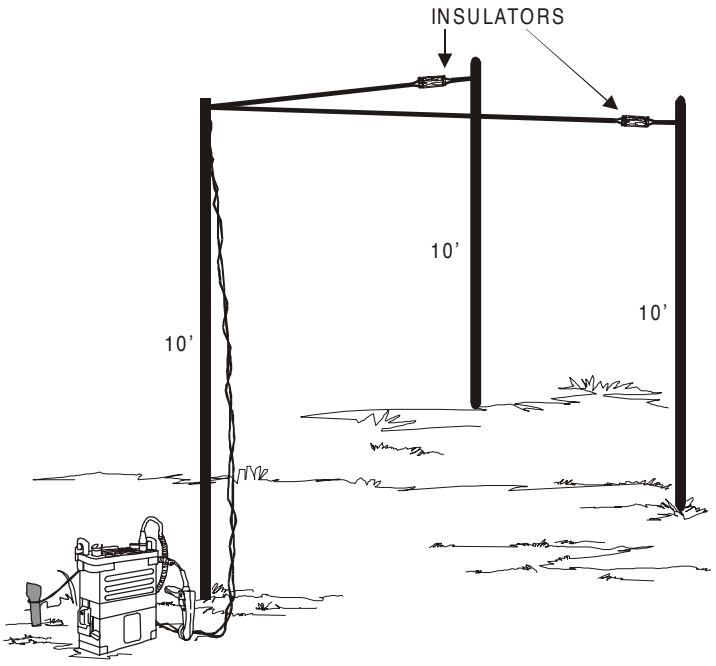

Figure O-12. Vee Antenna.

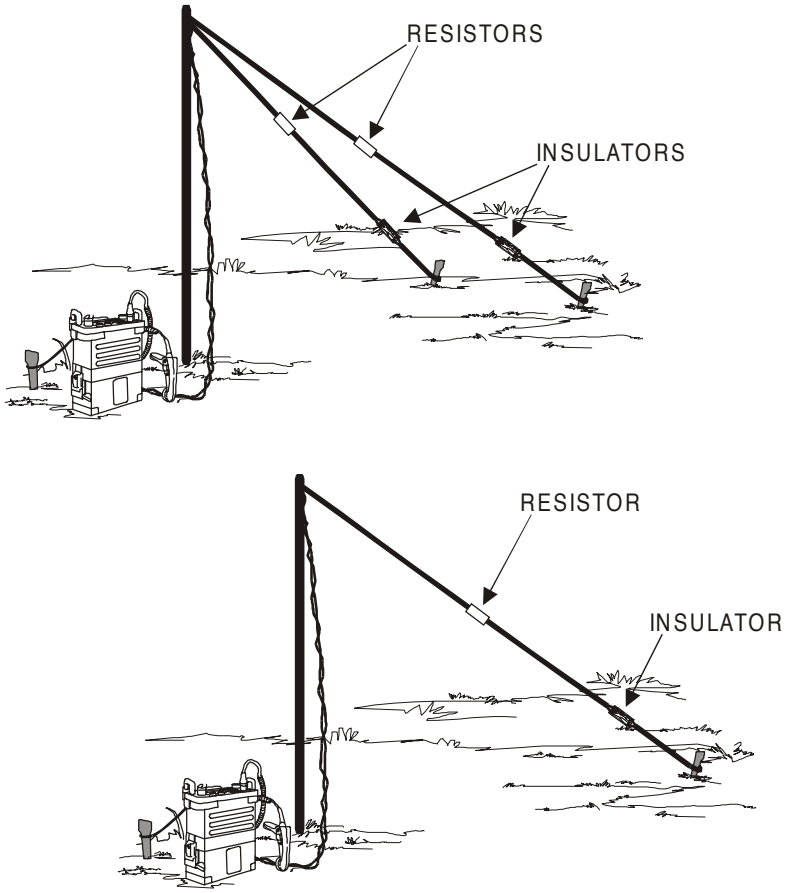

Figure O-13. Sloping Vee Antennas.

The resistors should be approximately 500 ohms and have a power rating at least one-half that of the output power of the transmitter being used. Without the resistors, the antenna radiates bidirectionally, both front and back.

The antenna must be fed by a balanced transmission line.

Use table O-1 to determine the angle and the length of the legs.

Table O-1. Leg Length for Vee Antennas.

Antenna Length (Wavelength)	Optimum Apex Angle (Degrees)
1	90
2	70
3	58
4	50
6	40
8	35
10	33

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Appendix P

Radio Operator's Checklist

Before you operate any radio set, get the appropriate equipment technical manual (TM) and carefully study the operating instructions. Refer to the panel diagrams, connections diagrams, and the paragraphs covering the description of components during the preliminary starting procedure. Make sure that the proper cables are connected to the proper panel connectors, and that the controls are correctly set. Even the most experienced operators should check their preliminary procedures against the TM references from time to time to insure accuracy and avoid damaging the equipment. Use the operational checklist and the equipment performance checklist to determine what to do to remedy any problems encountered during starting procedures and operation.

STEPS IN OPERATING RADIO SETS

Radio sets issued to a unit vary in type according to the communications requirements of the unit. For example, some sets may be completely contained in one assembly, while others may consist of separate components that must be properly connected to assemble a complete radio set. The following steps are generally required in operating a radio set.

Check the Set for Completeness

Make sure that all the necessary components and accessories are on hand and ready for use. Refer to the equipment basic issue items list in the TM. Never operate the transmitter without the antenna attached.

Inspect the Condition of the Knobs, Dials, Switches, and Controls

Look for knobs, dials, switches, and controls that are loose on their shafts, bind when being operated, won't operate, or are damaged in any other way. Make corrections where possible or report the faulty

condition to the CIS officer or CIS chief. Make sure that all knobs and exterior parts are on the set. Immediately report any that are missing.

Check the Condition of Plugs, Receptacles, and Connectors

Do not attempt to connect the set for operation until you are sure that the plugs and connectors are clean and in good condition, and the receptacles to which they must be connected are also clean and in good condition.

Check the Connections Diagrams

The connections diagrams in the equipment TM show the type and number of cables required to interconnect the components of the radio set for each type of operation. The radio set may be damaged if cables are connected to the wrong receptacles.

If the connectors don't match, it is possible to physically damage the pins or sleeves of the connector.

If a cable is connected to a receptacle into which it fits but does not belong, it may cause serious electrical damage to the equipment and, in some cases, injury to the operator.

Make Sure of Dial, Switch, and Control Settings

Some radio sets can be seriously damaged if the switches, dials, and controls are not set to the required initial settings before applying power or making the initial timing adjustments. Before applying power, check the equipment TM to be sure you performed all preliminary starting procedures. Be sure radios installed in vehicles are turned off before starting the vehicle's engine to avoid damage to radio equipment.

Follow the Starting Procedure

The equipment TM covers, in detail, the proper procedure for starting the radio set. If there is a specific sequence for starting the set, it is described in the manual. Perform the operations in the proper sequence.

Apply Power

After the proper connections are made, and all switches are properly set, power may be applied to the set.

Allow the Set to Warm Up

Radio sets usually require a warm-up period when first applying power in order to stabilize the equipment. In some cases, it is possible to damage a set by attempting to operate a set without allowing a warm-up period. Most sets are protected against such damage, but it is foolish to risk damage to a radio set by trying to put it on the air before it is ready.

Tune to the Desired Frequency (Channel)

Tune the transmitter to the frequency of the desired channel according to the procedures in the equipment TM. Use the methods that are given in the TM to check for correct tuning.

Check the Set for Normal Operation

While the set is in operation, check the indicators frequently to be sure that the set is operating correctly. If anything unusual occurs during operation, investigate it immediately. When necessary, turn off the power to the set and refer to the operational checklist and the equipment performance checklist in the equipment manual. If the corrections given in the operational checklist and the equipment performance checklist will not correct the trouble, report the condition to the unit electronics maintenance shop. Make sure that the condition of the set and the action taken are properly recorded on the maintenance records.

Use the Proper Procedure to Turn Off the Set

After operation (or if the set is being turned off because of improper operation) make sure that the controls, switches, and dials are properly set (this may not be required on some radios). Proceed to shut down the components of the set in the sequence specified in the equipment manual. Simple radios may require nothing more than turning the power

switch to its OFF position, but more complex sets may require elaborate shutdown procedures.

Operating Hints

Use a handset or headset, rather than a loudspeaker, if the incoming signal is weak. Make sure that the microphone or handset is in good condition. Speak directly into the microphone; speak slowly and distinctly.

Make sure that the vehicle's battery voltage (if radio set is vehicular-mounted) is within the correct range. Keep the engine running to charge the battery. Move the set or the vehicle, if necessary, to improve reception.

Lack of communications or poor communications may be caused by—

- Too great a distance between radio sets.
- Poor choice of location (siting) at one or both ends of the circuit.
- Terrain—hills or mountains.
- Noise and interference.
- Not enough transmitter power.
- Defective equipment.
- Improper adjustment of equipment.
- Ineffective antenna.
- Improper frequency assignment.

Poorly maintained equipment and improper operation can be just as effective in preventing communications as excessive distance or mountainous terrain. To avoid problems, observe the following precautions at all times:

- Study the TMs for the equipment you are using. They provide complete operating instructions and maintenance procedures.
- Keep your radio set clean and dry.
- Handle your radio set carefully.

Appendix Q

Glossary

Section I. Acronyms

AC ..... alternating current
AF ..... audio frequency
AM ..... amplitude modulation
ASAP ..... as soon as possible
ASIP ..... advanced systems improvement program

bps ..... bits per second

C4I ..... command, control, communications, computers, and intelligence
CDT ..... Central Daylight Time
CEOI. ..... communications-electronics operating instruction
CEP ..... circular error of probability
CIS ..... communications and information systems
cm ..... centimeter
COMINT ..... communications intelligence
COMSEC ..... communications security
CONUS ..... continental United States
COTS ..... commercial off-the-shelf equipment
CST ..... Central Standard Time

DACT ..... data automated communications terminal
DAMA ..... demand assigned multiple access
db ..... decibel
DC ..... direct current
DCT ..... digital communications terminal
dm ..... decimeter
dkm ..... dekameter
DMS ..... digital message system

DTE ..... data terminal equipment
DTG ..... date-time group

EA ..... electronic attack
EDT ..... Eastern Daylight Time
EEFI ..... essential elements of friendly information
EMCON ..... emission control
EMP ..... electromagnetic pulse
EP ..... electronic protection
EPLRS ..... enhanced position location reporting system
ES ..... electronic warfare support
EST ..... Eastern Standard Time
EW ..... electronic warfare

FAC ..... forward air controller
FLOT ..... forward line of own troops
FM ..... frequency modulation
FMF ..... Fleet Marine Force
FMFM ..... Fleet Marine Force manual

GHz ..... gigahertz
GPS ..... global positioning system

HF ..... high frequency
Hz ..... hertz
hm ..... hectometer

ID ..... identification
IED ..... imitative electromagnetic deception
IR ..... infrared

JC2WC ..... Joint Command and Control Warfare Center
JSC ..... Joint Spectrum Center

KHz ..... kilohertz
km ..... kilometer
kbps ..... kilobytes per second
kW ..... kilowatt

LOS ..... line of sight
LSB ..... lower side band
LUF ..... lowest usable frequency

m ..... meter
MAGTF ..... Marine air-ground task force
MCRP ..... Marine Corps reference publication
MCWP ..... Marine Corps warfighting publication
MDT ..... Mountain Daylight Time
MEU(SOC) ..... Marine expeditionary unit (special operations capable)
MFP ..... moisture fungusproofing paint
MHz ..... megahertz
mi ..... miles
MIJI ..... meaconing, intrusion, jamming, interference
mm ..... millimeter
MPF ..... maritime prepositioning force
MST ..... Mountain Standard Time
MUF ..... maximum usable frequency
MUX ..... multiplex

NBC ..... nuclear, biological, and chemical
NCS ..... net control station
NVIS ..... near vertical incidence skywave

ohm ..... a unit of electrical resistance (named for Georg Ohm)
OTH ..... over the horizon

prosign ..... precedure sign
proword ..... procedure word
PDT ..... Pacific Daylight Time
PLRS ..... position location reporting system
PST ..... Pacific Standard Time

RCU ..... remote control unit
RDF ..... radio direction finder
RF ..... radio frequency
RFI ..... radio frequency interference

SAA ..... satellite access authorization
SATCOM ..... satellite communications

SBB ..... switched backbone
SCR ..... single channel radio
SCSI ..... small computer systems interfaces
SHF ..... super-high frequency
SID ..... sudden ionospheric disturbance
SINCGARS ..... single-channel ground and airborne radio system
SIP ..... systems improvement program
SOP ..... standing operating procedure
SPEED ..... system planning, engineering, and evaluation device
SSB ..... single side band
SWR ..... standing-wave ratio

TACSAT ..... tactical satellite
TAMCN ..... table of authorized materiel control number
TCIM ..... tactical communications interface module
TDN ..... tactical data network
TEMPEST ..... an unclassified name referring to the means used to ensure computer security
TM ..... technical manual
TOD ..... time of day
TRANSEC ..... transmission security
TWA ..... tilt whip adaptor

UHF ..... ultrahigh frequency
USB ..... upper side band
UTC ..... Coordinated Universal Time
UTM ..... universal transverse mercator
UV ..... ultraviolet

VHF ..... very high frequency

WBA ..... whip-base adapter
WLC ..... whip loading coil
WWA ..... whip to wire adapter

yd ..... yard

Section II. Definitions

A

amplitude modulation (AM). Modulation in which the amplitude of the carrier wave is varied above and below its normal value in accordance with the intelligence of the signal being transmitted.

analog. A continuously variable signal which conveys information by the change of the value or magnitude of the signal. The signal can change in either amplitude, phase, frequency, or duration.

antenna gain. The effectiveness of a directional antenna as compared to a standard nondirection antenna. It is usually expressed as the ratio in decibels of standard antenna input power to directional antenna input power that will produce the same field strength in the desired direction. For a receiving antenna, the ratio of signal power values produced at the receiver input terminals is used. The more directional an antenna is, the higher is its gain.

authentication. A security measure designed to protect a communications system against acceptance of a fraudulent transmissions or simulation by establishing the validity of a transmission, message, or originator.

B

baseband. In a carrier (or subcarrier) wire or radio transmission system, the band of frequencies occupied by the signal before it modulates the carrier (or subcarrier) frequency to form the transmitted or radio signal.

black. The black designation is applied to all wire lines and equipment within a terminal or switching facility which handle encrypted traffic.

brevity code. A code which provides no security but which has as its sole purpose the shortening of messages rather than the concealment of their content.

bulk encryption. An application of on-line cryptographic operations where the encryption and decryption process is performed at designated points of technical interface within a communications system; input-output signals from various subscriber terminals thereby being crypto graphically processed at these points. The operation is performed in bulk; i.e., two or more channels processed simultaneously by one crypto security device.

C

call sign. Any combination of characters or pronounceable words which identifies a communication facility, a command, an authority, an activity, or a unit; used primarily for establishing and maintaining communications.

carrier. (1) The radio wave produced by a transmitter when there is no modulating signal, or any other wave, recurring series of pulses, or direct current capable of being modulated. Also called carrier wave. (2) A wave generated locally at a receiver that, when combined with the sidebands of a suppressed carrier transmission and a suitable detector, produces the modulating wave.

communications security (COMSEC). The protection resulting from all measures designed to deny unauthorized persons information of value which might be derived from the possession and study of telecommunications, or to mislead unauthorized persons in their interpretation of the results of such possession and study. Communications security includes cryptosecurity, transmission security, emission security, and physical security of communications security materials and information.

critical frequency. The highest frequency at which a given wave at any given time will, if transmitted vertically, be refracted back to earth by a layer of the ionosphere.

cryptography. The art or science which pertains to the various means and methods for rendering plain text unintelligible, and reconverting unintelligible texts into intelligible language; application of that science by means other than cryptanalysis.

D

data. Representation of facts, concepts, or instructions in a formalized manner suitable for communication, interpretation, or processing by humans or by automatic means. Any representations such as characters or analog quantities to which meaning is or might be assigned.

date-time group (DTG). The date and time, expressed in digits and zone suffix, the message was prepared for transmission. (Expressed as six digits followed by the zone suffix; first pair of digits denotes the date, second pair the hours, third pair the minutes.)

decibel (dB). A unit used to express the magnitude of a change in signal or sound level. A change of three decibels is the change in power level of a pure sine wave that is just barely detectable by the human ear. The difference in decibels between two signals is 10 times the common logarithm of their ratio of powers or 20 times the common logarithm of their ratio of voltages or currents. One decibel is one-tenth of a bel.

digital. A signal having discrete states, usually two, such as the presence or absence of a voltage. The signal is given meaning by assigning numerical values or other information to the various possible combinations of the discrete states of the signal.

directed net. A net in which no station other than the net control station can communicate with any other station, except for the transmission of urgent messages, without first obtaining the permission of the net control station.

E

electromagnetic radiation. Radiation made up of oscillating electric and magnetic fields and propagated with the speed of light.

electronic warfare (EW). Any military action involving the use of electromagnetic and directed energy to control the electromagnetic spectrum or to attack the enemy. The three major subdivisions within electronic

warfare are: electronic attack, electronic protection, and electronic support measures.

F

free net. A net in which any station may communicate with any other station in the same net without first obtaining permission from the net control station to do so.

frequency. The number of complete cycles per unit of time for a periodic quantity such as alternating current, sound waves, or vibrating objects. Frequency is expressed in hertz, kilohertz, megahertz, and gigahertz.

frequency band. A continuous range of frequencies extending between two limiting frequencies.

Frequency hopping. A method of jumping from frequency to frequency in synchronization with one another in a random order at a rate of up to 100 times per second. Frequency hopping is the preferred method of communication with SINCGARS radios.

frequency modulation (FM). Frequency modulation is the process of varying the frequency (rather than the amplitude) of the carrier signal in accordance with the variations of the modulating signals. The amplitude or power of the FM carrier does not vary during modulation. A frequency modulation system is practically immune to atmospheric and man-made interference.

frequency spectrum designation. VLF (very low frequency): below 30 kHz (0.03 MHz). LF (low frequency): 30-300 kHz (0.03-0.3 MHz). MF (medium frequency): 300-3000 kHz (0.3-3 MHz). HF (high frequency): 3-30 MHz. VHF (very high frequency): 30-300 MHz. UHF (ultra high frequency): 300-3000 MHz. SHF (super high frequency): 3000-30,000 MHz (3-30 GHz). EHF (extremely high frequency): 30-300 GHz.

G

ground wave. A radio wave that is propagated over the earth and is ordinarily affected by the presence of the ground and the troposphere. The ground wave includes all of the components of a radio wave over the earth except ionospheric and tropospheric waves. The ground wave is refracted because of variations in the dielectric constant of the troposphere, including the condition known as a surface duct.

guard. To maintain a continuous receiver watch with transmitter ready for immediate use. A complete log is to be kept.

H

half-duplex—Refers to a mode of transmission in which communication between two terminals occurs in either direction, but in only one direction at a time. This is the typical mode of operation for tactical single-channel radios.

hertz (Hz). A unit of frequency equal to one cycle per second.

I

imitative communication deception (ICD). The introduction of radiation into enemy systems which imitate the enemy's emissions.

in the clear. In plain text. When cryptographic devices are not used to protect a transmitted signal.

ionosphere. A region in the earth's outer atmosphere where ions and electrons are present in quantities sufficient to affect the propagation of HF radio waves. It begins about 30 miles above the earth and extends above 250 miles, with the height depending on the season of year and the time of day. The chief layers of the ionosphere and their approximate heights are:
D layer—30 to 60 miles
E layer—60 to 90 miles
F layer—90 to 250 miles

J

jamming (electromagnetic). The deliberate radiation, reradiation, or reflection of electromagnetic energy for the purpose of preventing or reducing an enemy's effective use of the electromagnetic spectrum, and with the intent of degrading or neutralizing the enemy's combat capability.

L

line of sight (LOS). The straight unobstructed path between two points.

link. In communications, a general term used to indicate the existence of communications facilities between two points.

log. A chronological record of station events.

N

needline. A requirement to establish communications between two units or agencies.

net (communications). An organization of stations capable of direct communications on a common channel or frequency.

net call sign. A call sign which represents all stations within a net.

net control station. A communications station designated to control traffic and enforce circuit discipline within a given net. Also called NCS.

O

obstacle gain. The increase in signal strength obtained over a long radio communications path where a mountain obstacle or range of hills is located about halfway between transmitting and receiving antennas. This obstacle gain offsets some of the path losses normally expected.

operating signals. Three letter groups used as necessary in connections with operations or communications to convey orders, instructions, requests, reports, and information to facilitate communications.

P

precedence. In communications, a designation assigned to a message by the originator to indicate to communications personnel the relative order of handling and to the addressee the order in which the message is to be noted.

procedure sign (prosign). One or more letters or characters, or combination thereof, used to facilitate communications by conveying in a condensed standard form certain frequently used orders, instructions, requests, and information related to communications.

procedure word (proword). A word or phrase limited to radio telephone procedure used to facilitate communication by conveying information in a condensed standard form.

Propagation. A phenomenon by which any wave moves from one point to another.

pulse code modulation (PCM). The form of modulation in which the modulating signal is sampled, and the sample quantitized and coded so that each element or information consists of different kinds and/or numbers of pulses and spaces.

pulse position modulation (PPM). A form of pulse modulation in which intelligence is conveyed by varying the time interval by which successive pulses are displayed from their normal times of occurrence.

push to talk operation. Voice communications on a circuit in one direction at a time in which operation of a switch is required prior to and during transmission.

R

radio frequency (RF). A frequency in which coherent electromagnetic radiation of energy is useful for communications purposes. The useful range is from approximately 10 kilohertz to 100,000 megahertz.

radio silence. A condition in which all or certain radio equipment capable of radiation is kept inoperative. (In combined or U.S. joint or intra Service communications the frequency bands and/or types of equipment affected will be specified.)

red. Designation applied to transmission lines, equipment, systems, and/or areas passing unencrypted signals.

S

satellite communications. Use of communication satellites, passive reflecting belts of dipoles or needles, or reflecting orbiting balloons to extend the range of radio communications by returning signals to earth from the orbiting objective, with or without amplification.

simplex. Refers to a mode of operation in which communication between two terminals can take place in only one direction.

skip distance. The minimum separation at which radio waves over a specified frequency can be transmitted at a specific time between two points on the earth by reflection from the regular ionized layers of the ionosphere.

skip zone. A region, in relation to a given transmitter, in which no signal would be predicted, either due to direct repetition or due to reflected waves.

sky wave. A radio wave that reaches the receiving location after refraction from the ionosphere.

T

tactical radio net. A functional radio net used by a commander for immediate and direct control of the fire and maneuver or movement of his subordinate units.

traffic. All transmitted and received messages.

transceiver. A radio transmitter and receiver combined in one unit and having switching arrangements such as to permit use of one or more circuit components for both transmitting and receiving.

transducer. A device that transfers or changes one type of energy into another form. An example is a loudspeaker, which changes electrical energy into acoustic (mechanical) energy.

Appendix R

References and Related Publications

Joint Publication (Joint Pub)

1-02	Department of Defense Dictionary of Military and Associated Terms

Marine Corps Doctrinal Publication (MCDP)

6	Command and Control

Marine Corps Warfighting Publications (MCWPs)

6-2	MAGTF Command and Control Operations (under development)
6-22	Communications and Information Systems

Marine Corps Reference Publications (MCRPs)

6-22A	TALK II-SINCGARS Multiservice Communications Procedures for the Single-Channel Ground and Airborne Radio System
6-22B	Multiservice Procedures for Spectrum Management in a Joint Environment
6-22D	Antenna Handbook (under development)

Army Field Manuals (FMs)

11-32	Combat Net Radio Operations
24-11	Tactical Satellite Communications
24-18	Tactical Single-Channel Radio Communications Techniques
24-19	Radio Operator's Handbook

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Ultra-Compact Ku band Rectenna

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Abstract — This paper addresses an innovative and ultra-compact rectenna designed for energy harvesting or wireless power transfer applications. The presented rectenna uses a printed cross dipoles antenna array and a rectifier implemented with only one silicon Schottky diode. Experimental results show that 1.15 mW of DC power can be obtained for an optimal load impedance of 500 Ω using a compact rectenna (2.5 cm^2 or 0.6 square wavelength) illuminated by an electric field of 60 V/m at 14.7 GHz.

Index Terms — Rectennas, microwave energy harvesting, wireless power transfer

I. Introduction

Microwave spectrum (beyond 10 GHz) presents an increased interest for wireless power transfer [1] or energy harvesting [2] applications. Rectenna topologies working at such frequencies were designed in the past [2]-[7] by using exclusively GaAs Schottky diodes. One of the main challenges of rectenna design, especially at such high frequencies, is to provide a high-efficiency by using a compact design and low-cost electronic components (e.g., silicon Schottky diodes). This paper addresses an innovative rectenna topology and the associated design methodology. The topology and the associated design rules are presented in Section II. The experimental results reported in Section III demonstrate that this topology allows implementing an ultra-compact and high efficiency microwave rectenna.

II. Rectenna Topology and Design

The proposed topology selected to meet the two main design goals (i.e., conversion efficiency and structure compactness) is shown in Fig. 1. It is composed by: (i) a compact antenna array of two crossed printed dipoles located on the top of the PCB; (ii) a rectifier using only one Schottky diode. The rectifier (except the Schottky diode) is located at the bottom side of the PCB. A ground plane can be positioned below antenna/rectenna to immunize the rectenna performances from the electromagnetic properties of any mechanical support, to increase the gain of the antenna and consequently, to enhance the overall efficiency of the rectenna.

A. Antenna design

Three criteria were taken into account during antenna design and optimization: (i) *compactness*: the total size occupied by the antenna should be as small as possible, (ii) *high gain/high efficiency*: for maximizing the amount of the RF power available at the input of the rectifier and, (iii) *input impedance matching*: antenna input impedance (Z_{inA}) has to match the input impedance of the rectifier (Z_{inR}) for maximizing the power transfer between antenna and rectifier ($Z_{inA}=Z_{inR}^*$, where Z_{inR}^* denotes the complex conjugate of Z_{inR}) at the targeted operating frequency.

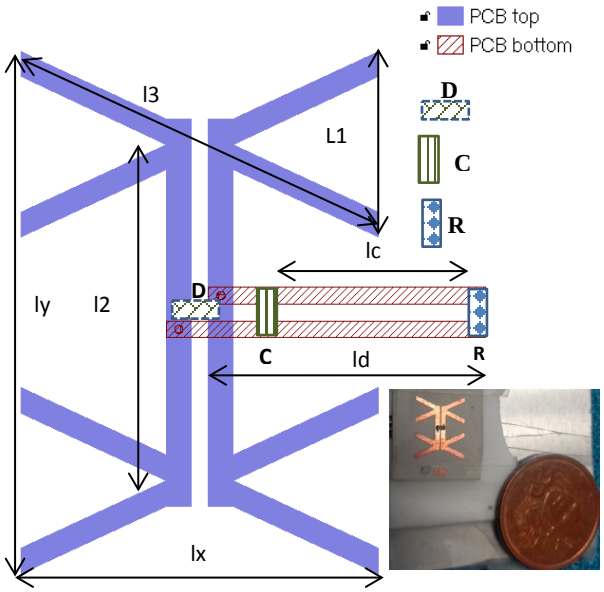

Fig. 1. Top view (not to scale) of the layout of the rectenna, its main geometrical dimensions and a photo (inset, right corner) of the manufactured rectenna

B. Rectifier design

A low-cost silicon Schottky diode (SMS201 from Aeroflex Metelics) in a molded plastic (DFN) package was adopted for this design. This diode can be used for broadband zero bias detectors or power detection up to 10 dBm (frequency below 26.5 GHz). The SMS201 diode has a good thermal behavior and can be mounted using a classical soldering process at high temperature (260°C) while the GaAs diodes require a more sensitive mounting process at a lower temperature (< 150°C). A RF shunt capacitor is connected between the diode and the load resistor as shown in Fig. 1. The simulation model of the rectifier was implemented into AWR software (Fig. 2). The diode was modeled based on the Metelics application note [8]. The rectenna was designed and fabricated on Rogers 6002

substrate (relative permittivity: 2.94, loss tangent: 0.0012 and thickness: 508 µm).

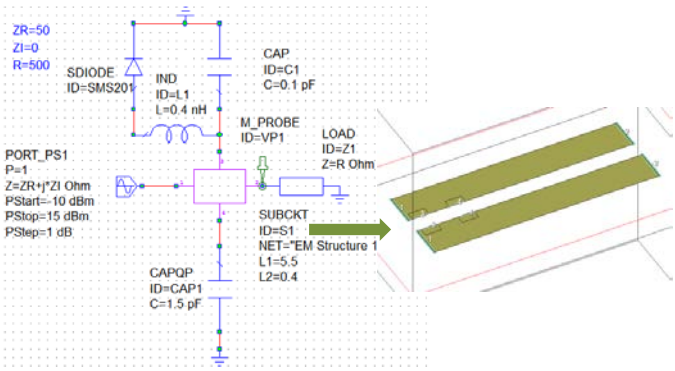

Fig. 2. Simulation model (AWR) of the rectifier.

C. Design methodology and optimization process

Rectenna design, simulation and optimization at such a high frequency involve the use of full-wave electromagnetic simulation tools combined with non-linear (e.g. harmonic balance) electrical circuit models.

First, non-linear simulations were performed by using the AWR model reported on Fig. 2 and an optimal value was determined for the shunt RF capacitor (C=1.5 pF) and for the load resistance (R=500 Ω). The coplanar stripline sections supporting the rectifier were modeled as a sub-circuit at the electromagnetic level due to the lack of appropriate transmission line models in AWR. The via-hole used to connect the diode (mounted on the top side of the PCB) with the bottom of the PCB (the rectifier, except the diode, is located on the bottom side of the PCB) was not simulated in our model. A parametric analysis was performed in order to find the best position for the capacitor mounting. At this stage of the simulation it was found that the distance between the diode and the capacitor (ld-lc) should be as small as possible.

Second, a simulation model of the antenna array (top side of the PCB) was performed using Feko electromagnetic software. The main geometrical parameters of antennas are: the length of the crossed dipole arms (ld), the angle between crossed dipoles (α), the array step (l2), the strip width (w) and, the gap between the strips of the coplanar stripline (g). The length of the crossed dipole arms (ld) and the array step (l2) must be close to the half-wavelength at the operating frequency. In order to increase the gain of the antenna array and to immunize the rectenna performances from the electromagnetic properties of any mechanical support a metallic ground plane was positioned bellow the PCB. At this step of the design the radiation pattern and the current distribution on the strip section of antenna array were checked to verify the proper radiation mechanism. The input of antenna array is loaded by the rectifier impedance.

The rectenna layout is very compact and the antenna performances can be impacted by the presence of the rectifier. Thus an improved simulation model was developed by taking into account the entire layout (top and bottom side of the PCB). The diode was replaced by a voltage port while the capacitance (1.5 pF) and the load impedance (500 Ω) were modeled as port loads. By using this simulation model the impact on antenna performances of the overall rectifier layout (except the non-linear behavior of the diode) was taken into account.

Finally the antenna layout (l3 and l2) as well as the position of the shunt RF capacitor were tuned in order to ensure the matching condition $Z_{inA}=Z_{inR}^{*}$.

III. Results

The layout of the optimized rectenna was accommodated with the manufacturing tolerances available in a University unit equipped for general (low frequency) PCB manufacturing. A metallic plane was positioned at 1.2mm below the rectenna with a 1.2 mm thick Rohacell (dielectric constant in the range of 1.08) intercalated as spacer between rectenna and its reflector. The main dimensions of the manufactured rectenna (see Fig. 1) are: lx=8.4mm, ly=12.4mm, l1=4.4mm, l2=8mm, ld=6.5mm, lc=5.5 mm l3=10 mm, w=0.6 mm and g=0.4 mm. The overall surface of the PCB is 2.5 cm^2 ($0.6\lambda^2$).

A. Experimental setup

An experimental setup (shown in Fig. 3) was used in order to recreate the electromagnetic environment existing on antenna panels of the broadcasting satellites. A microwave signal generated from an Anritsu MG3694B generator was injected at the input of a horn antenna which illuminated the rectenna under test with a linear polarized E-field.

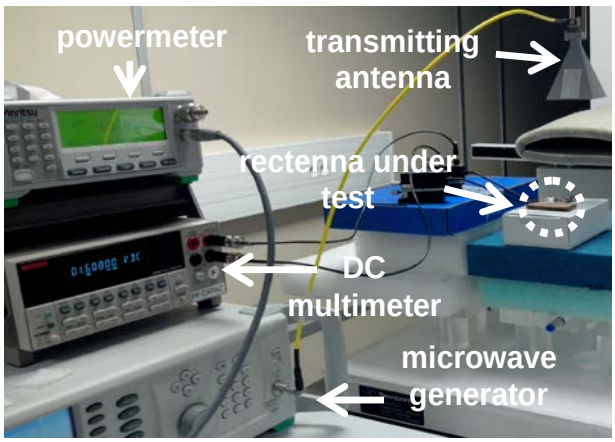

Fig. 3. Experimental setup.

An automatic acquisition routine was implemented in Labview software from National Instruments to speed-up the acquisition process. The harvested DC voltage was measured by using a DC multimeter. The DC power can be computed from the measured DC voltage as long as the load is known. The RF output power delivered by the signal generator was 24 dBm and the measured loss due to the coaxial cable and connectors between antenna and the signal generator was in the range of 2.5 dB in the operating frequency band.

B. Experimental and simulation results

The simulated radiation pattern (gain) at 14.7 GHz is shown in Fig. 4. A standard 3D Cartesian coordinate system with the vertical Oz axis perpendicular to the rectenna/antenna surface is chosen here.

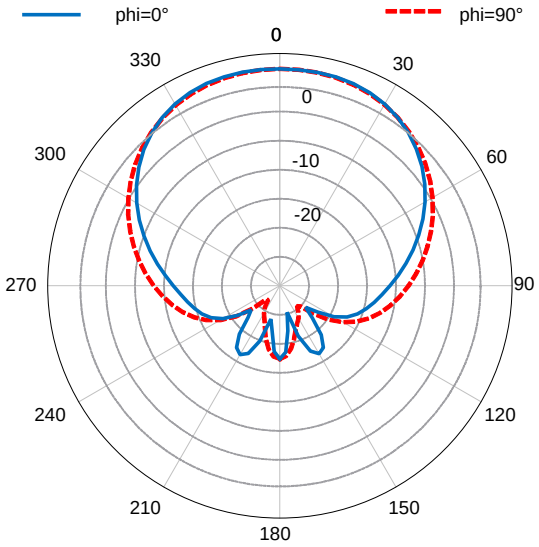

Fig. 4. Simulated (FEKO) gain of the antenna array: (xoz-plane: phi=0° and yoz-plane: phi=90°) at the frequency f=14.7 GHz.

The DC voltage on the input of a variable load (from 0.1 kΩ to 10 kΩ) was measured for the fabricated rectenna. Fig. 5 shows the DC power measured with a load impedance of 500Ω. The rectenna delivers the maximum power of 1.15 mW (load: 500 Ω) at 14.7 GHz. Fig. 6 shows the measured harvested DC power as function of the load impedance.

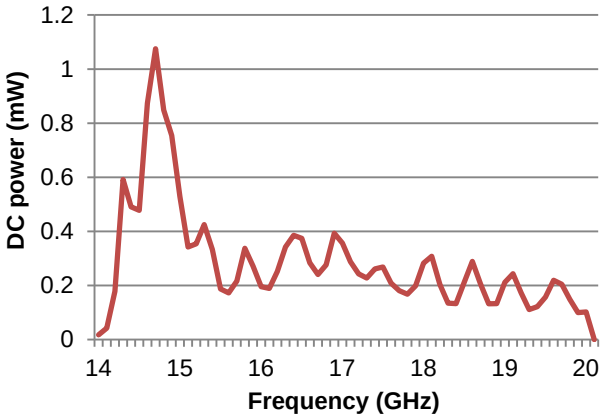

Fig. 5. Measured DC power on a load of 500 Ω as function of frequency.

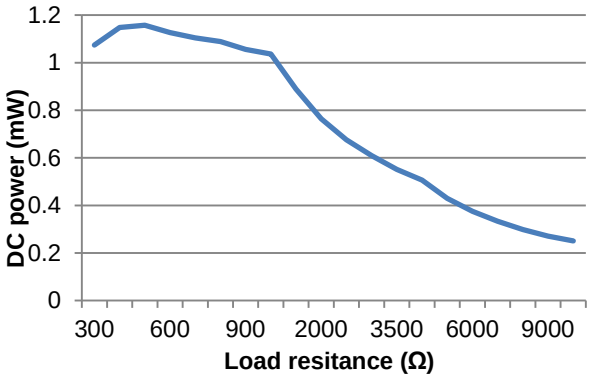

Fig. 6. Measured DC power at 14.7 GHz as function of the load resistance.

The efficiency η (in %) of the rectenna can be computed by using the following definition [9]:

$$\eta_1 = \frac{P_{DC}}{S \cdot A_G} \cdot 100 \qquad (1)$$

$$\eta_2 = \frac{P_{DC}}{S \cdot A_{eff}} \cdot 100 = \frac{4 \cdot \pi \cdot P_{DC}}{S \cdot G_R \cdot \lambda^2} \cdot 100 \qquad (2)$$

where P_{DC} is the harvested DC power, S is the incident electromagnetic power density, A_G (in cm^2) denotes the area of the radiating surface, A_{eff} is the antenna effective area, G_R is the gain of the (rectenna's) antenna and λ is the wavelength of the illuminating electromagnetic wave. The efficiency η_1 can be viewed as a 'worst-case' definition because $A_{eff} \leq A_G$ for any passive antenna. The power density ($\mu W/cm^2$) can be computed as a function of E-field effective value E (V/m) on the antenna surface or as a function of the RF power P_t delivered to the transmitting horn antenna of gain G_t and positioned at the distance *d* from the rectenna, as follows:

$$S = \frac{E^2}{120 \cdot \pi} \cdot 100 = \frac{30 \cdot P_t \cdot G_t}{d^2 \cdot 120 \cdot \pi} \cdot 100 \qquad (2)$$

Here $P_t \approx 21.5$ dBm, $G_t \approx 16.9$ dBi and d=19 cm. Thus E≈60 V/m and S~955 $\mu W/cm^2$. By taken into account the overall area of the rectenna $A_G \approx 2.5$ cm^2 a conversion efficiency around of η_1=48% (η_2=66% by taking into account the simulated gain of G_R=7.4 dBi depicted in Fig. 4) at 14.7 GHz (DC power of 1.15 mW) were obtained. A comparison with the state-of-the-art (operating frequency beyond 10 GHz) is presented in Table I (note that all the diodes -except SMS201- are GaAs diode). Few papers present the measured efficiency of the overall rectenna and, most papers report only the efficiency of the rectifier block of the rectenna.

TABLE I

COMPARISON WITH PUBLISHED RECTENNAS (BEYOND OF 10 GHZ).

f (GHz)	η (%)	Diode Reference	Size (mm^2)	Reference & Comment
35	35	M/A Com MA4E-1317	>110	[4]: efficiency (received RF power/output DC power) measured using a free space setup for S=30 mW/cm^2 and a load of 50Ω
24	54.6	M/A Com MADS-001317-1320AG	NR*	[5]: efficiency (received RF power/output DC power) measured using a near field setup for 130 mW of RF input power at the input of the rectifier and a load of 400Ω
25.7	17	M/A Com MA4E-2502L	328.5	[7]: only the efficiency of the rectifier part is reported; value obtained for a RF power of 8 dBm at the diode input and a load of 1500Ω
18.8	42	MZBD-9161	280	[2]: E-field of 91 V/m (S=2.2 mW/cm^2) for a load of 510Ω
24	24	M/A Com MA4E-1317	>1800	[6]: S= 10 mW/cm^2; use of a circular polarized antenna array
14.7	**48 (66)**	**SMS201**	**250**	**This paper: E=60 V/m; S=0.95 mW/cm^2 for a load of 500Ω.**
23.15	25	SMS201	250	[10]: E=73 V/m (S=1.4 mW/cm^2) for a load impedance of 300 Ω

* NR : Not Reported

IV. CONCLUSION

A compact rectenna ($0.6\lambda^2$) operating in Ku-band was designed and characterized. Despite of the use of a low-cost silicon Schottky diode the manufactured rectenna exhibits an efficiency of 48% at 14.7 GHz. This rectenna exhibits an excellent trade-off between compactness and efficiency.

ACKNOWLEDGMENT

This work was partially supported by CNES French Space Agency within the framework of R&T n° 115052 contracts. The electronic laboratory of University of Toulouse (Paul Sabatier) is acknowledged for PCB manufacturing.

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[4] Y.-J. Ren, M.-Y. Li, K. Chang, "35 GHz rectifying antenna for wireless power transmission," *IET Electronics Letters*, vol.43, no.11, pp.602-603, May 24 2007.

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[8] http://www.aeroflex.com/AMS/Metelics/pdfiles/SMS201.pdf

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[10] A. Takacs, H. Aubert, A. Luca, S. Charlot, S. Fredon, L. Despoisse, "Rectenna Design for K Band Application", *2014 European Microwave Conference (EUMC'2014)*, Rome, Italy, 5-10 Oct'2014.

RF circuit design: Basics

Akira Matsuzawa

Tokyo Institute of Technology

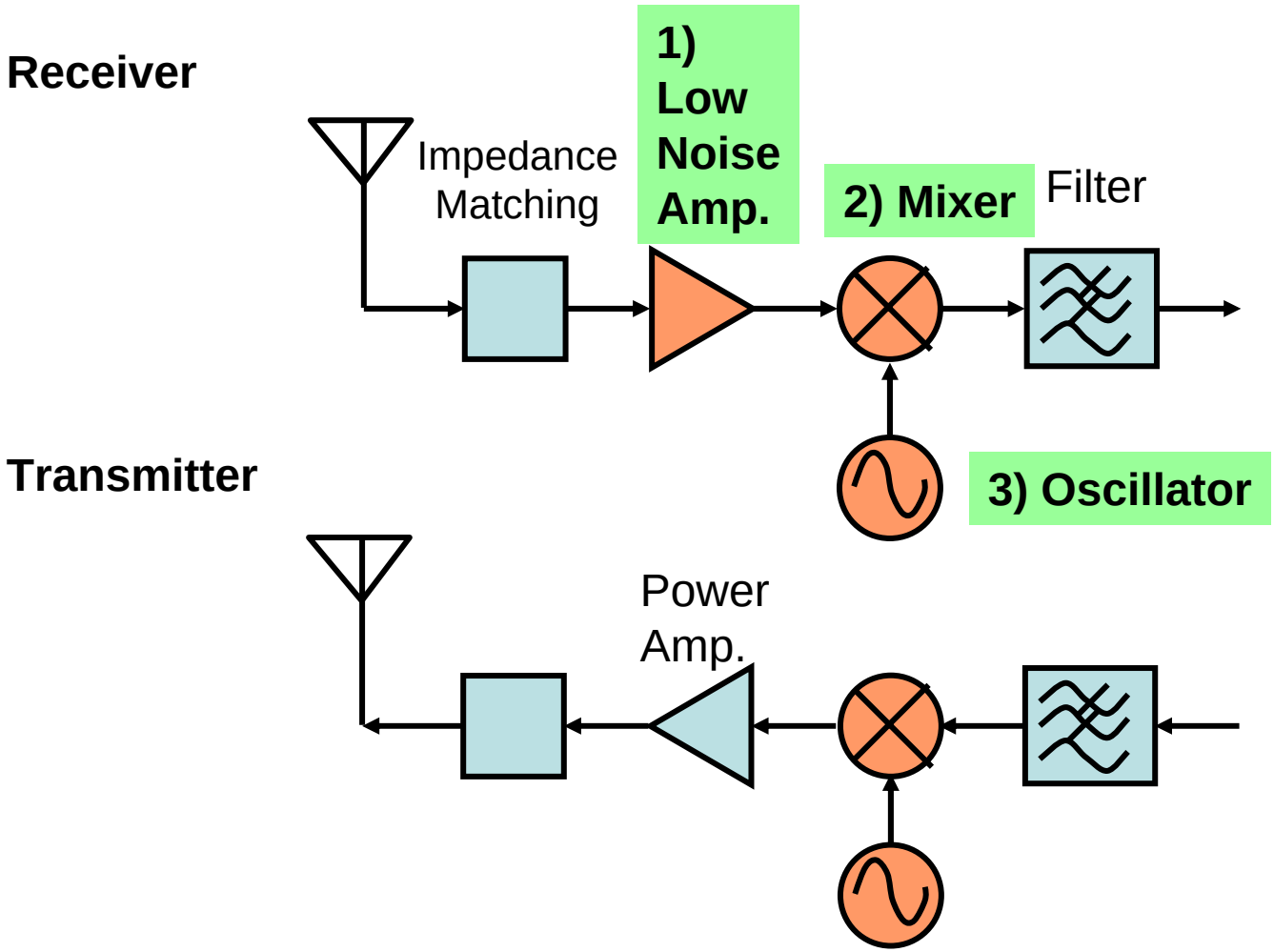
Contents

- Building blocks in RF system and basic performances
- Device characteristics in RF application
- Low noise amplifier design
- Mixer design
- Oscillator design

Basic RF circuit block

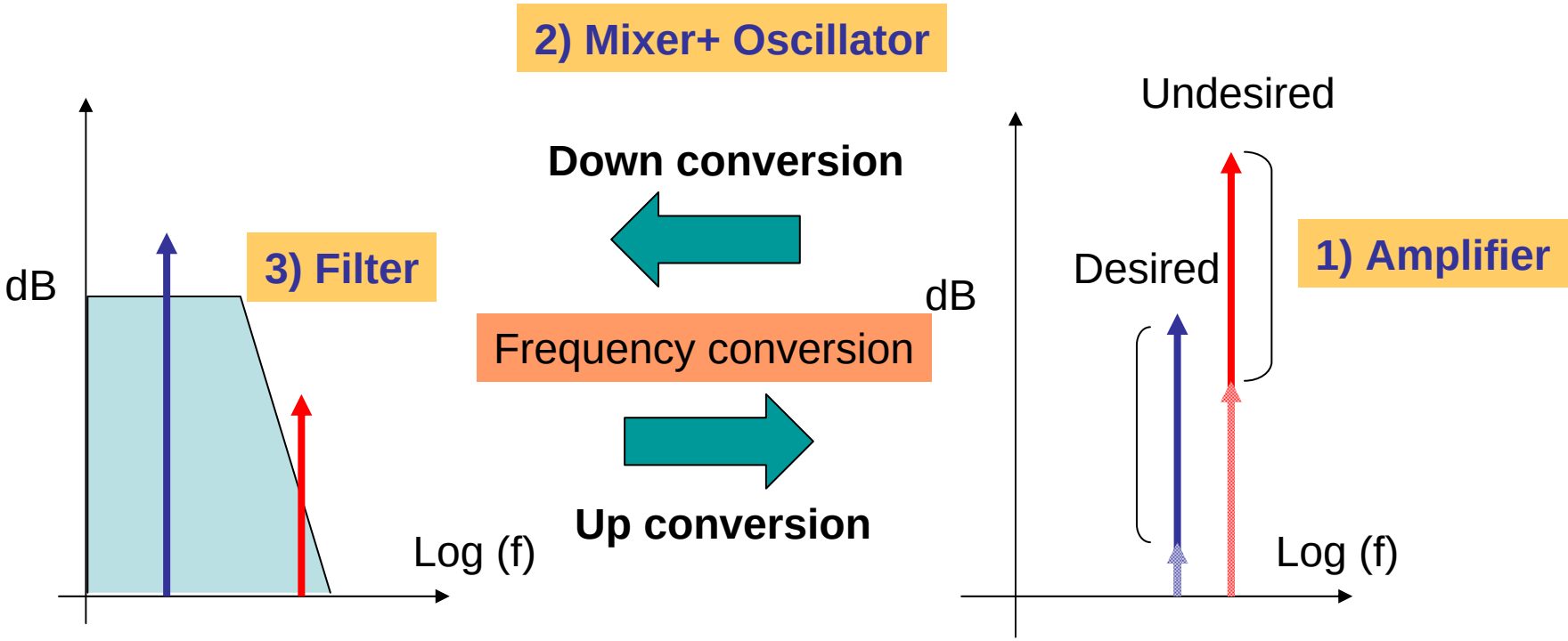
RF systems are composed of limited circuits blocks.

LNA, Mixer, and Oscillator will be discussed in my talk.



Basic functions of RF building blocks

Amplifier, frequency converter (mixer +oscillator), and filer are basic function blocks in RF system.



RF Amplifier

- **Gain:** Amplify small signal or generate large signal.
- **Noise:** Smaller noise and larger SNR.
- **Linearity:** Smaller non-linearity.

Non-linearity generates undesired frequency components.

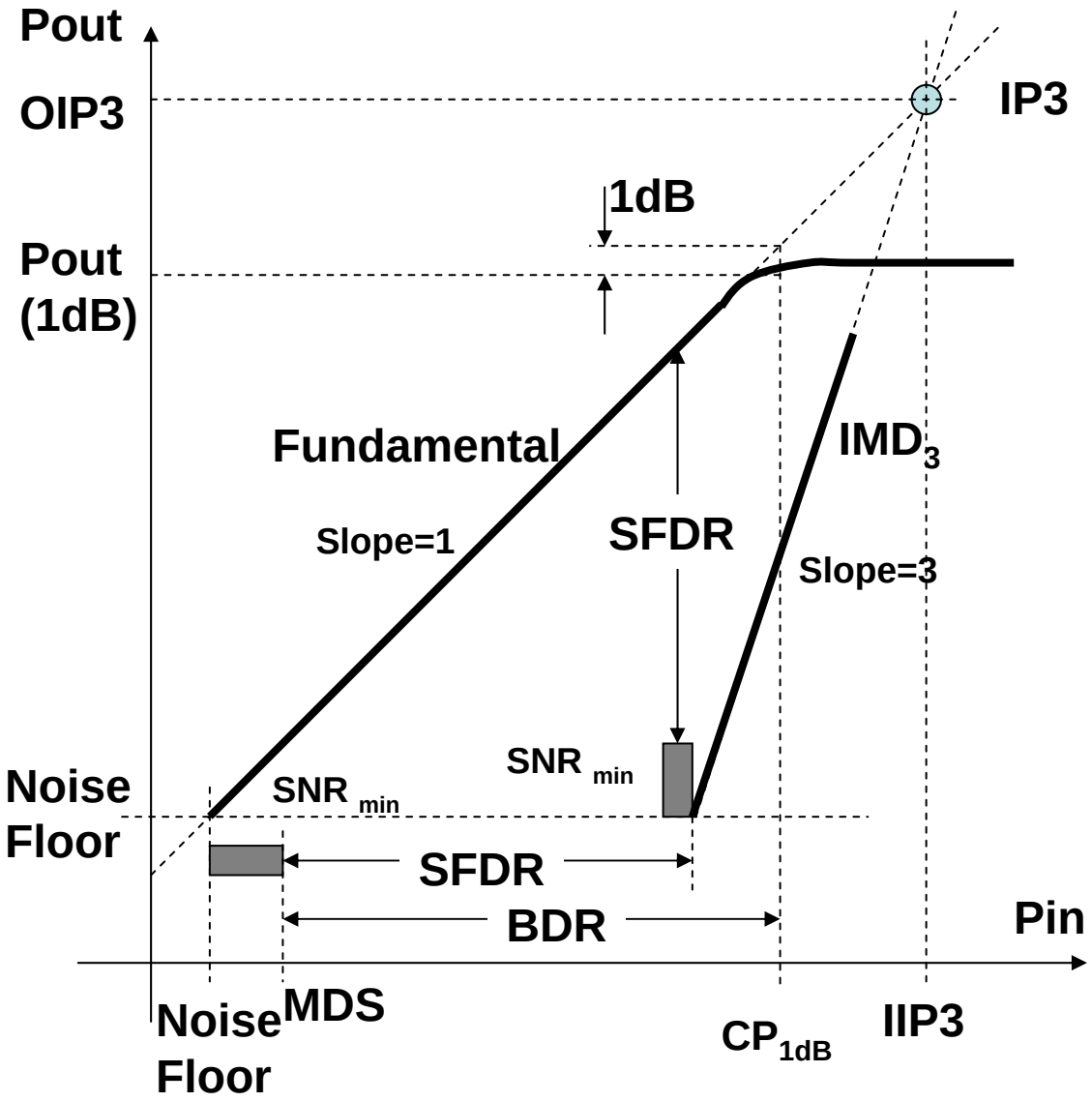
$$v_{out}(t) = \alpha_1 v_{in}(t) + \alpha_2 v_{in}^2(t) + \alpha_3 v_{in}^3(t) + ...$$

$$\left(\cos(\omega_1 t) + \cos(\omega_2 t)\right)^2 = 2 + \cos(2\omega_1 t) + \cos(2\omega_2 t) + \cos((\omega_1 - \omega_2)t) + \cos((\omega_1 + \omega_2)t)$$

$$\left(\cos(\omega_1 t) + \cos(\omega_2 t)\right)^3 = \frac{1}{2}\cos((2\omega_1 - \omega_2)t) + \frac{1}{2}\cos((2\omega_2 - \omega_1)t) + ...$$

Input and output characteristics

Distortion and noise are important factors in RF amplifier, as well as power and gain.



Dynamic range

$$Noise\ Floor = \underbrace{-174dBm}_{\text{kT limitation}} + NF + 10\log\underbrace{BW}_{\text{Bandwidth}}$$

SFDR: Spurious free dynamic range

The input power range over which third order inter-modulation products are below the minimum detectable signal level.

$$SFDR = \frac{2}{3}\left(IIP_3 - Noise\ Floor\right) - SNR_{\min}$$

BDR: Blocking dynamic range

$$BDR = P_{1dB} - Noise\ Floor - SNR_{\min}$$

MDS: Minimum detectable signal level= Noise Floor +SNR_{min}

Non-linearity

CP_{1dB}: The input level at which the small signal gain has dropped by 1dB.

$$CP_{1dB} = \sqrt{0.145\left|\frac{\alpha_1}{\alpha_3}\right|}$$

IMD3: The third order inter modulation term

IP3: The metric third order intercept point. It is the point where the amplitude of third order inter modulation is equal to the that of fundamental.

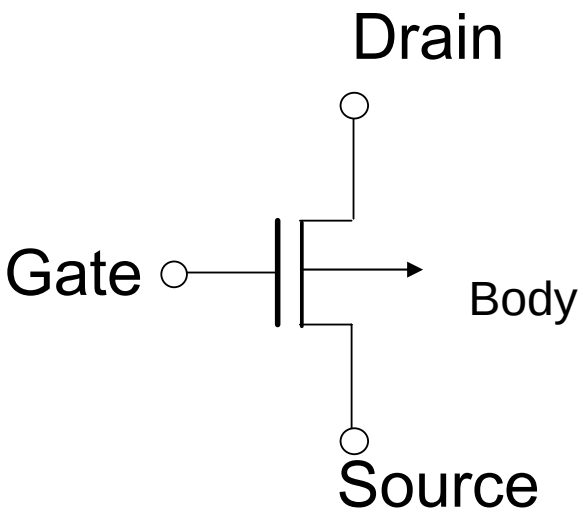
$$A_{IP3} = \sqrt{\frac{4}{3}\left|\frac{\alpha_1}{\alpha_3}\right|}$$

IIP3: Input referred intercept point
OIP3: Output referred intercept point

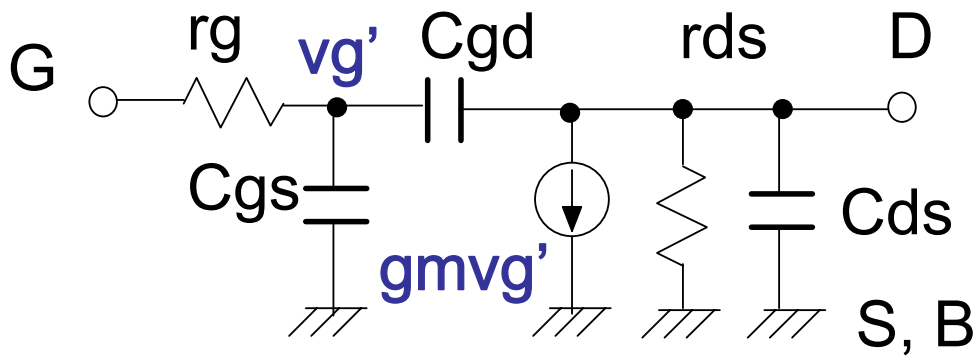
$$P_{out} - IMD_3 = 2 \cdot \left(IIP_3 - P_{in}\right)$$

MOS transistor

Intrinsic gate voltage and gm are the most important factors in RF CMOS.



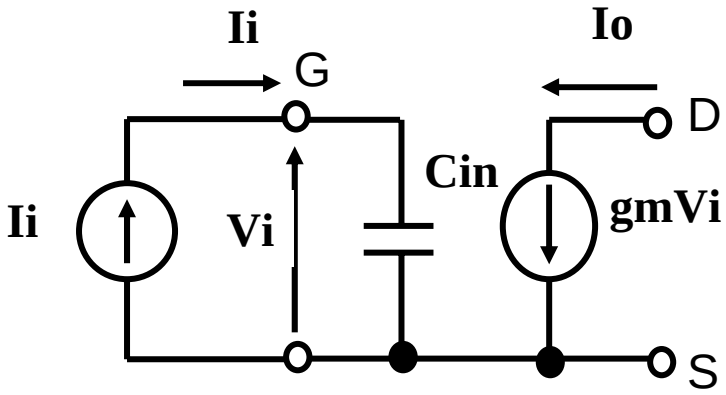
MOS Transistor



Equivalent Circuit

Cutoff frequency: f_T

For higher f_T, increase gm and decrease Cin.



f_T: Frequency at which the current gain is unity.

$$Ii = Iio\sin(\omega t)$$ Input current

$$Vi = \frac{Iio}{\omega C_{in}}\cos(\omega t)$$ Gate voltage

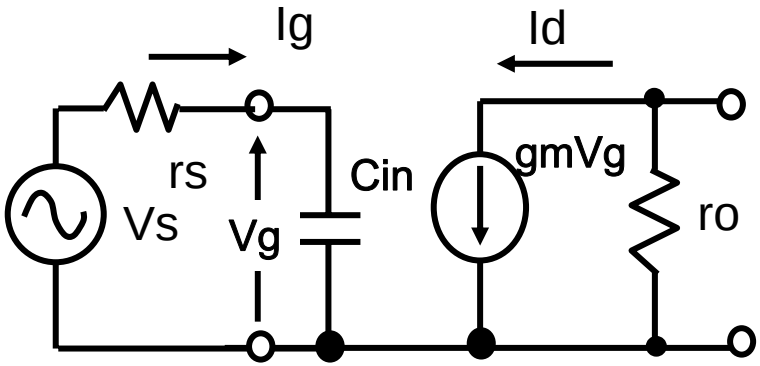
$$Io = gmVi = \frac{gmIio\cos(\omega t)}{\omega C_{in}}$$ Output current

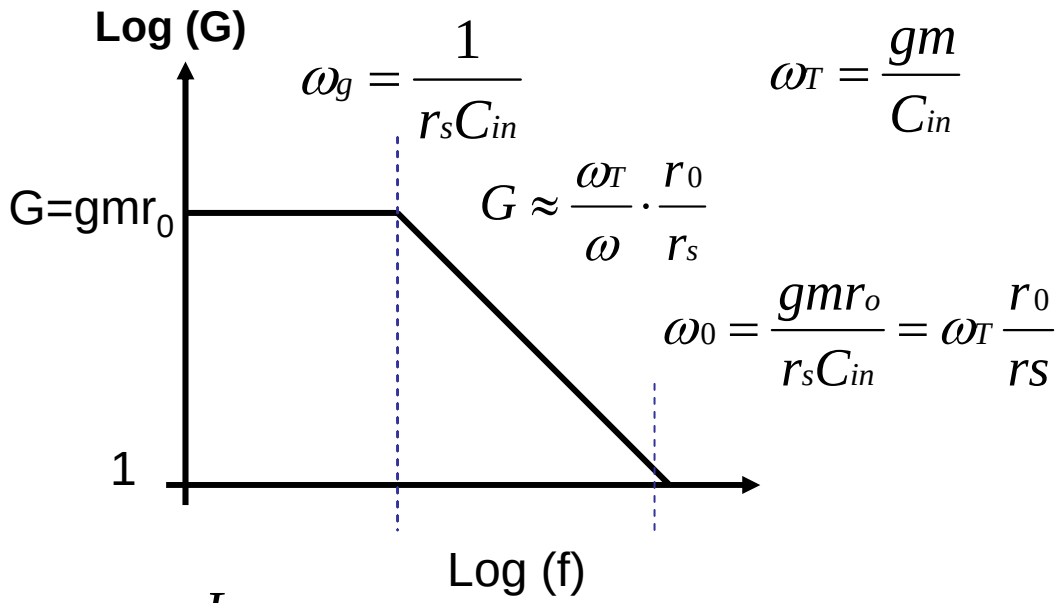
$$\therefore f_T = \frac{gm}{2\pi Cin}$$

Proportional to gm
Inversely proportional to Cin

Amplifier gain

For higher voltage gain, increase gm, f_T, ro (Q), and decrease input and gate resistance





$$\omega_g = \frac{1}{r_s C_{in}} \qquad \omega_T = \frac{gm}{C_{in}}$$

$$G \approx \frac{\omega_T}{\omega} \cdot \frac{r_0}{r_s}$$

$$\omega_0 = \frac{gmr_o}{r_s C_{in}} = \omega_T \frac{r_0}{rs}$$

For the larger gain

Fundamentally larger gmr_0 $G \approx gmr_o \approx \dfrac{I_{ds}}{\left(\dfrac{V_{eff}}{2}\right)} \cdot r_o$ ⇨ Larger I_{ds} or r_o

Larger Q

Higher f_T and lower r_s

V_{eff} is difficult to reduce $\quad \because r_o = Q\omega_0 L = \dfrac{Q}{\omega_0 C}$

→ Distortion and C_{in} increase

Characteristics of gm (Basic)

Gm is proportional to the Ids and inversely proportional to the V_{eff}.

V_{eff} is proportional to square root of I_{ds} and inversely proportional to square root of (W/L) ratio.

Square law region

$$I_{ds} = \frac{\mu C_{OX}}{2n}\left(\frac{W}{L}\right)(V_{gs} - V_T)^2 = \frac{\mu C_{OX}}{2n}\left(\frac{W}{L}\right)V_{eff}{}^2$$

$$gm \equiv \frac{dI_{ds}}{dV_{gs}} = \frac{\mu C_{OX}}{n}\left(\frac{W}{L}\right)V_{eff}$$

$$gm = \sqrt{\frac{2\mu C_{OX}}{n}\left(\frac{W}{L}\right)I_{ds}}$$

$$gm = \frac{Ids}{\left(\frac{V_{eff}}{2}\right)}, \quad \frac{gm}{Ids} = \frac{1}{\left(\frac{V_{eff}}{2}\right)}$$

$$V_{eff} = \sqrt{\frac{2n}{\mu}\cdot\frac{1}{C_{ox}}\cdot\frac{L}{W}\cdot I_{ds}}$$

$$V_{eff} \propto \sqrt{L\frac{Ids}{W}} = \sqrt{L \cdot J_{ds}}$$

Scaling W/L ratio

V_{eff} is proportional to square root of drain current density.

Non-ideal effects to square low region

At larger V_{eff} and lower V_{eff}, two non-ideal effects are not negligible .

Low V_{eff} Sub-threshold region

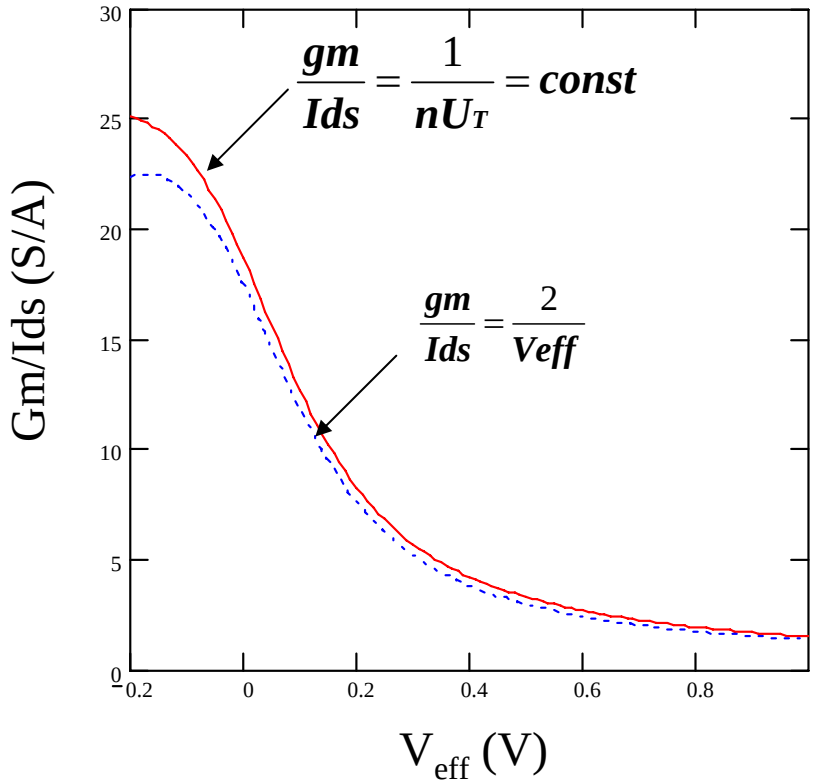
(Weak inversion)

$$I_{ds} = I_{so}\exp\left(\frac{V_{gs}}{nU_T}\right)$$

$$gm = \frac{I_{ds}}{nU_T} \qquad \frac{gm}{I_{dS}} = \frac{1}{nU_T}$$

High V_{eff} Mobility degradation

$$\mu \approx \frac{\mu_0}{1+\theta V_{eff}}, \quad \theta \approx \theta_0 + \frac{\mu_0}{v_c L}$$



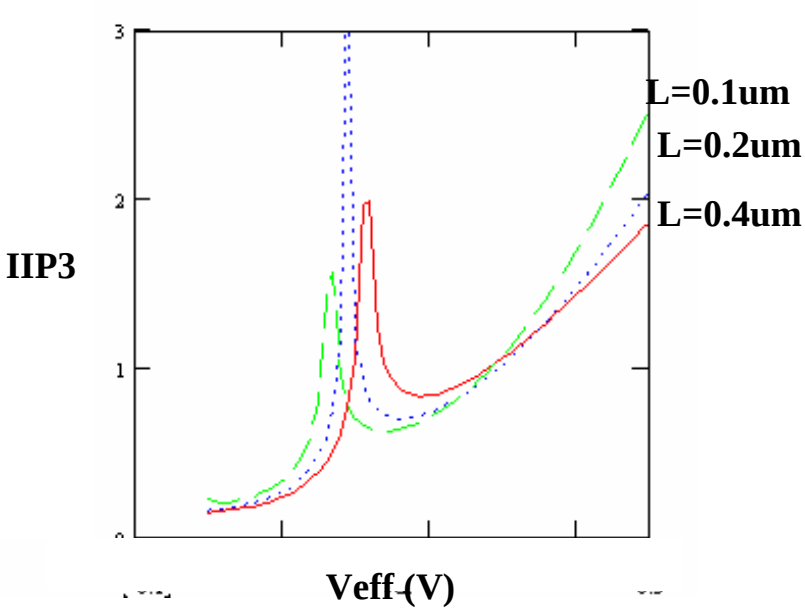
This effect becomes larger at large V_{eff} and short channel length.

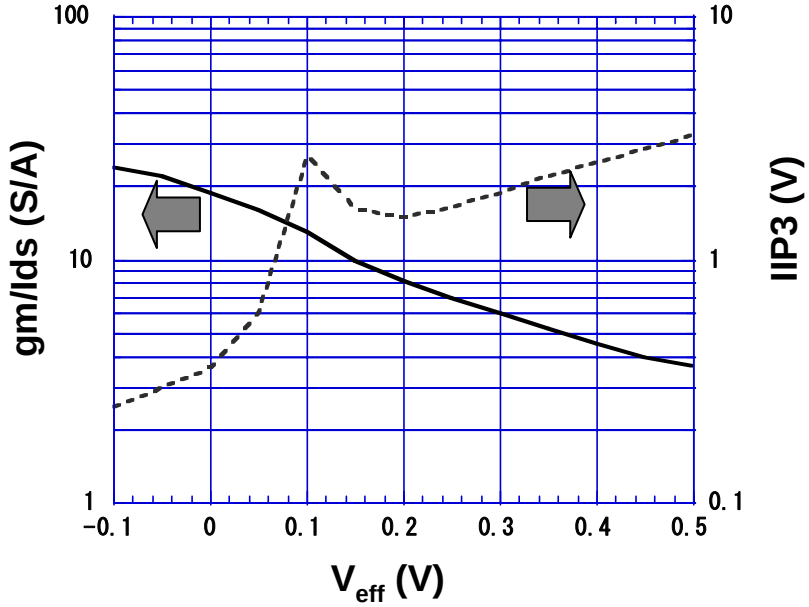
Distortion

Lower V_{eff} gives higher gm, bur results in higher distortion.
To obtain lower distortion (higher IIP3), we must increase V_{eff}.

Higher gm and lower distortion means higher Ids.

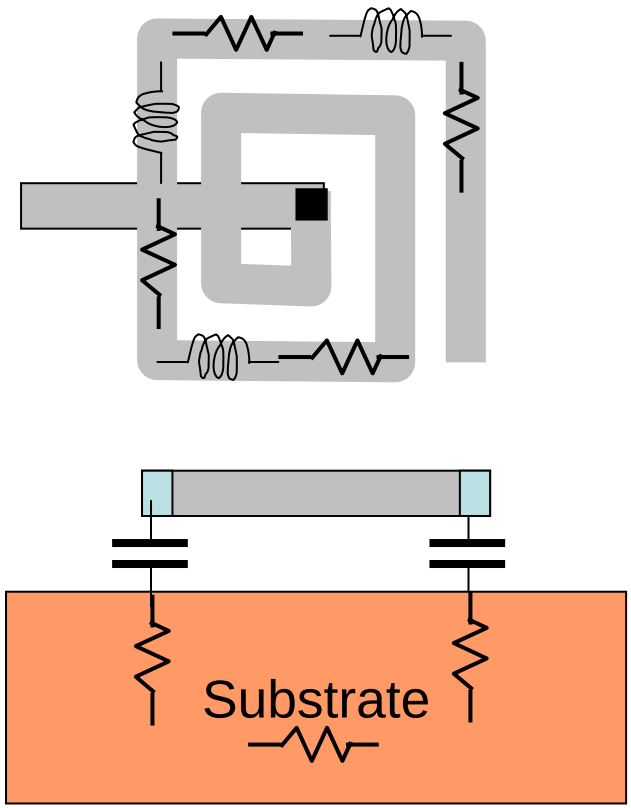
$$Ids = a_1 V_{eff} + a_2 V_{eff}^{\ 2} + a_3 V_{eff}^{\ 3} + \cdots \qquad a_3 \equiv \frac{1}{6}\frac{d^3 I_{ds}}{dV_{eff}^{\ 3}} \qquad I_{IP3} = \sqrt{\frac{4}{3}\left|\frac{a_1}{a_3}\right|}$$

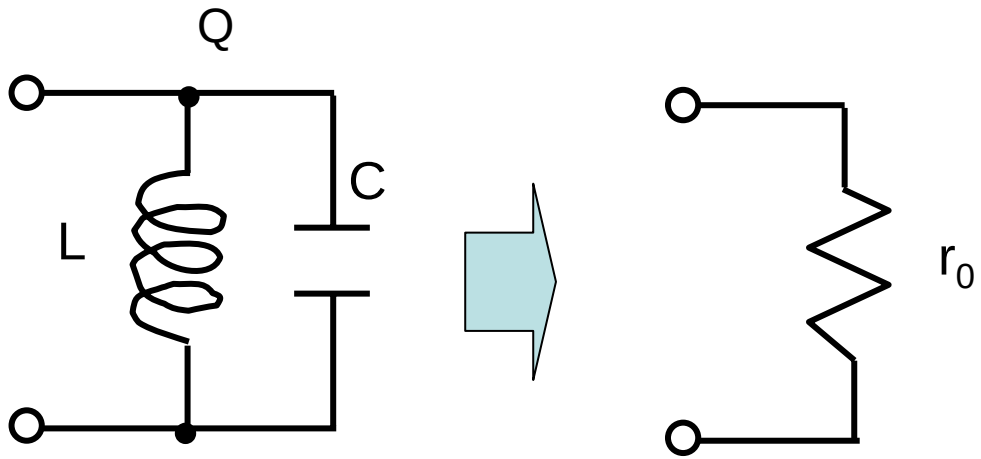




LC resonator

LC resonator can be regarded as resistance at the resonance frequency.





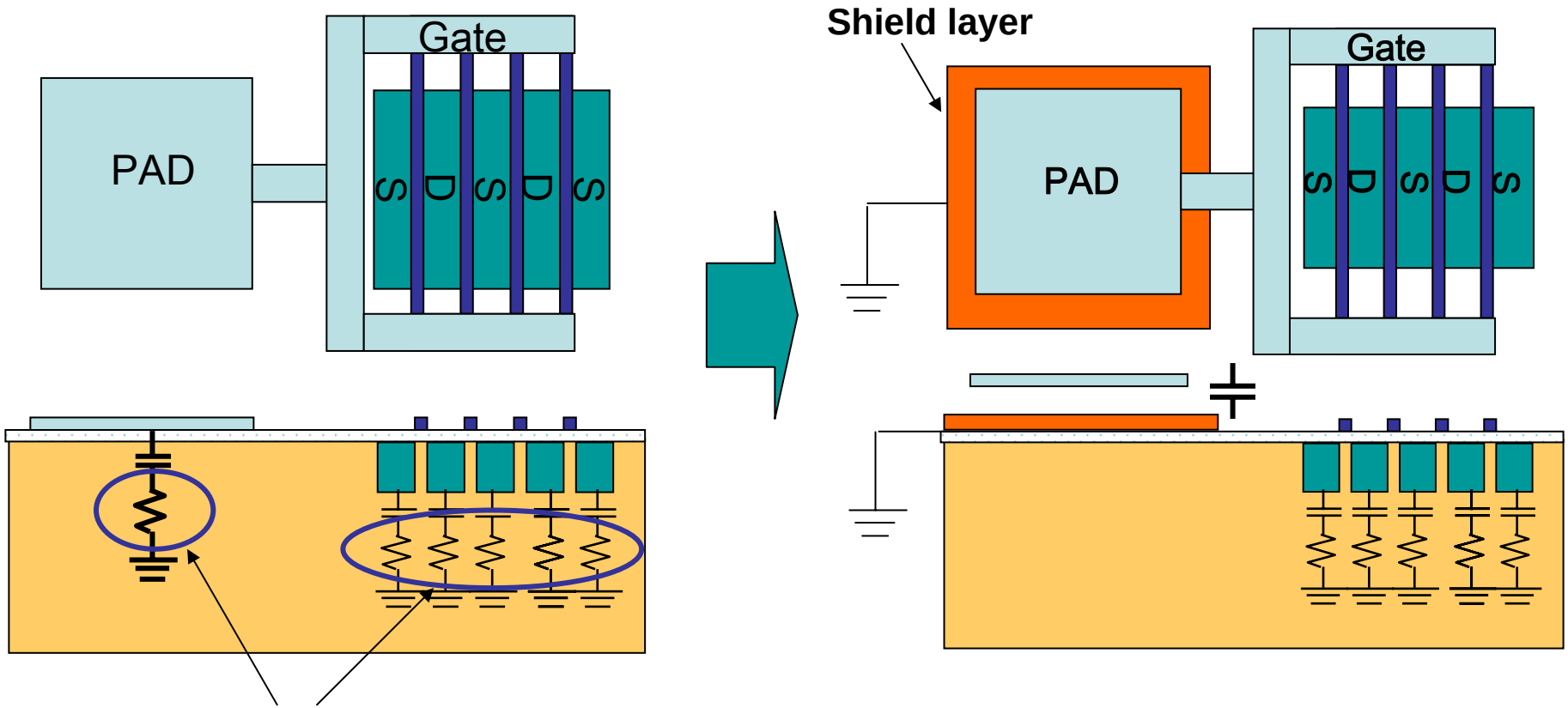
$$\omega_0 = \frac{1}{\sqrt{LC}} \qquad r_o = Q\omega_0 L = \frac{Q}{\omega_0 C}$$

Substrate effect

Substrate should be treated as resistive network.

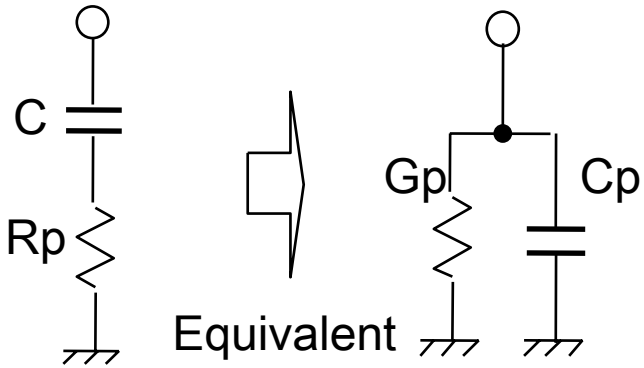
This substrate resistance causes RF power loss and noise generation.

Shielding can reduce this effect.



Power loss in substrate

Very low resistance or high resistance realizes low power loss.

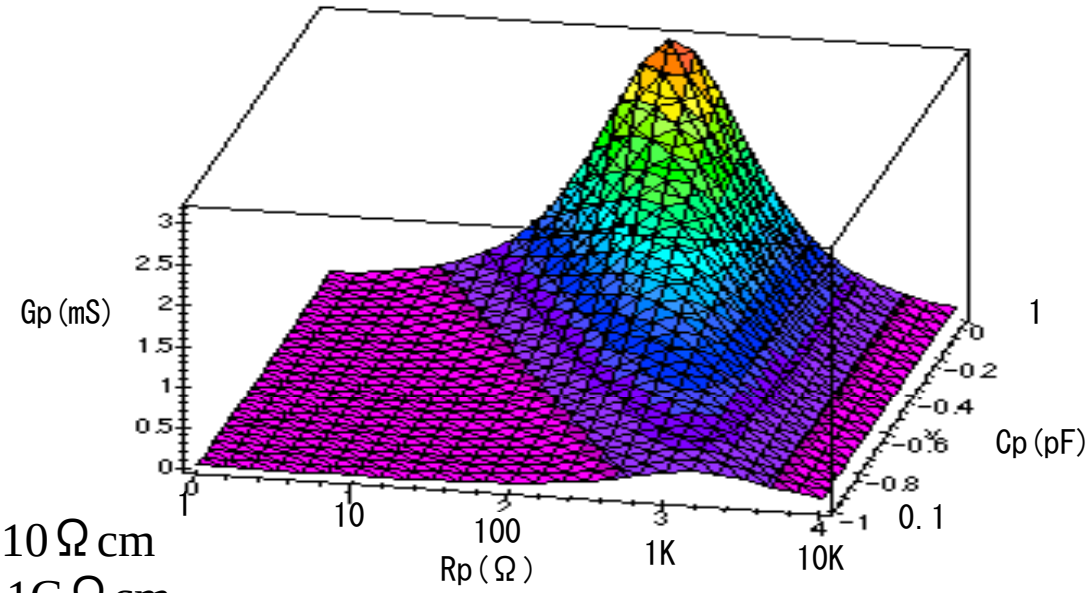


Higher C and moderate R_{sub} results in higher power loss.

$$G_p = \frac{1}{R_p} \cdot \frac{\left(\frac{\omega}{\varpi_p}\right)^2}{1 + \left(\frac{\omega}{\varpi_p}\right)^2}$$

$$C_p = C \frac{1}{1 + \left(\frac{\omega}{\varpi_p}\right)^2}$$

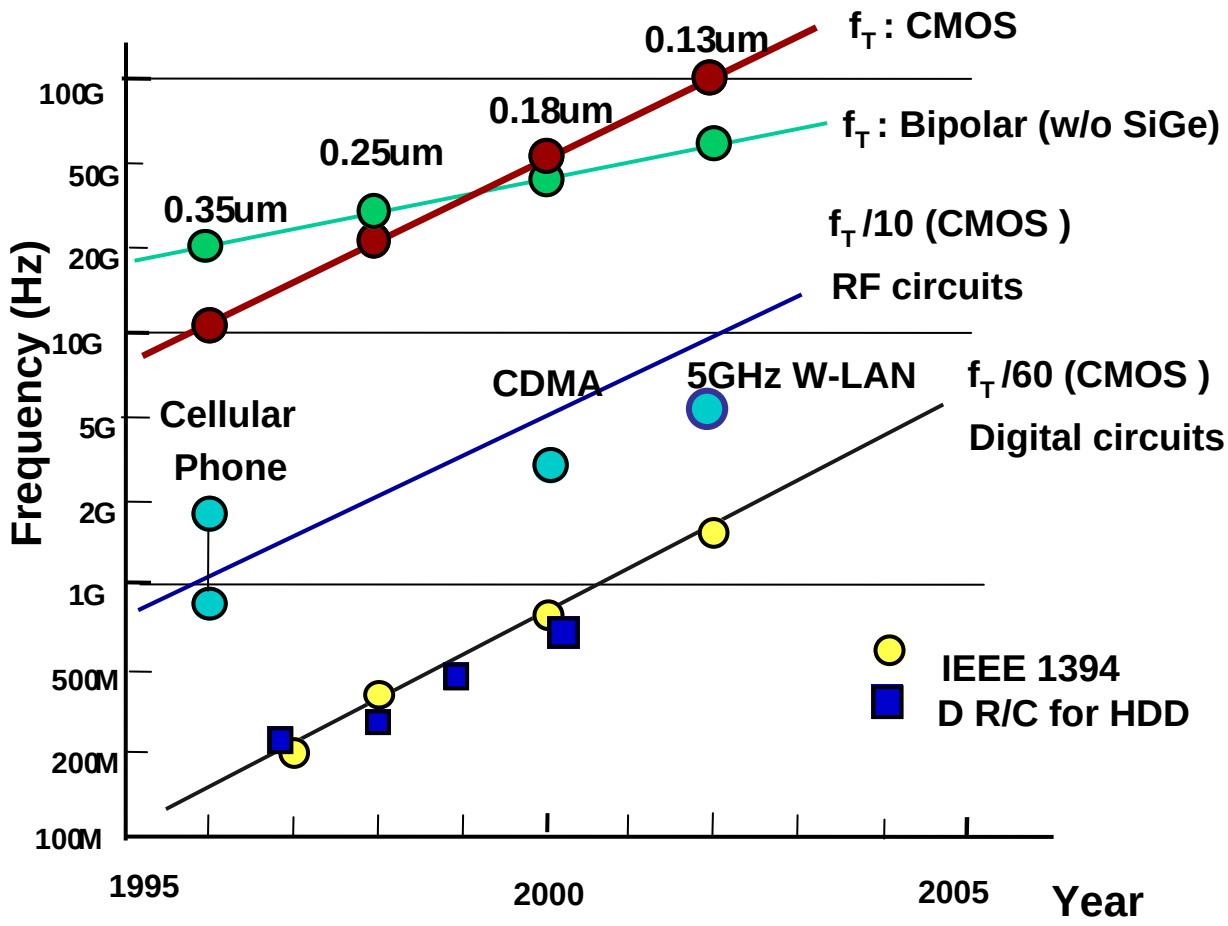
$$\varpi_p = \frac{1}{R_p C}$$



MOS: 10 Ω cm
GaAs: 1G Ω cm

GHz operation by CMOS

The cutoff frequency of MOS becomes higher than that of Bipolar. Over several GHz operations have attained in CMOS technology



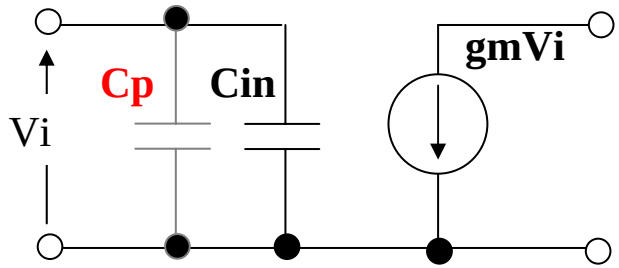
$$f_T \equiv \frac{gm}{2\pi C_{in}}$$

$$f_{Tpeak} \approx \frac{v_{sat}}{2\pi L_{eff}}$$

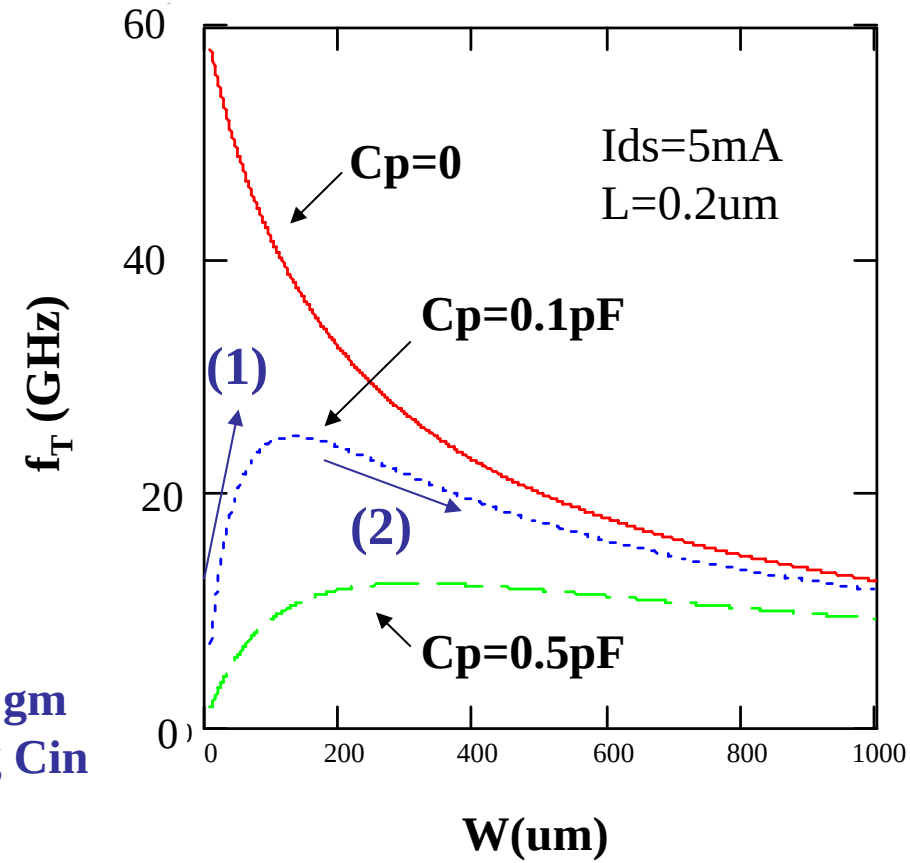
Effect of parasitic capacitance to f_T

f_T of actual circuit is reduced by a parasitic capacitance.
There is an optimum gate width to obtain highest f_T.

$$f_T \equiv \frac{gm}{2\pi(C_{gs}+C_{gd}+C_p)}$$



$C_{in}=C_{gs}+C_{gd}$



Region(1); Increased by increasing gm
Region(2); Decreased by increasing Cin

f_T: MOS vs. Bipolar

Even if f_T of MOS is the same as that of Bipolar,
f_T of MOS is easily lowered by a parasitic capacitance.
Because, gm of MOS is ½ to ¼ of that of Bipolar at the same current.

Small parasitic capacitance is a key for RF CMOS design.

MOS

$$gm \equiv \frac{Ids}{\left(\frac{V_{eff}}{2}\right)}$$

$$V_{eff\,\min} = 2nU_T \quad \text{n: 1.4}$$

Veff/2: 50-100mV (actual ckt.)

$$f_T \equiv \frac{gm}{2\pi Cin}$$

Bipolar

$$gm \equiv \frac{Ic}{U_T}$$

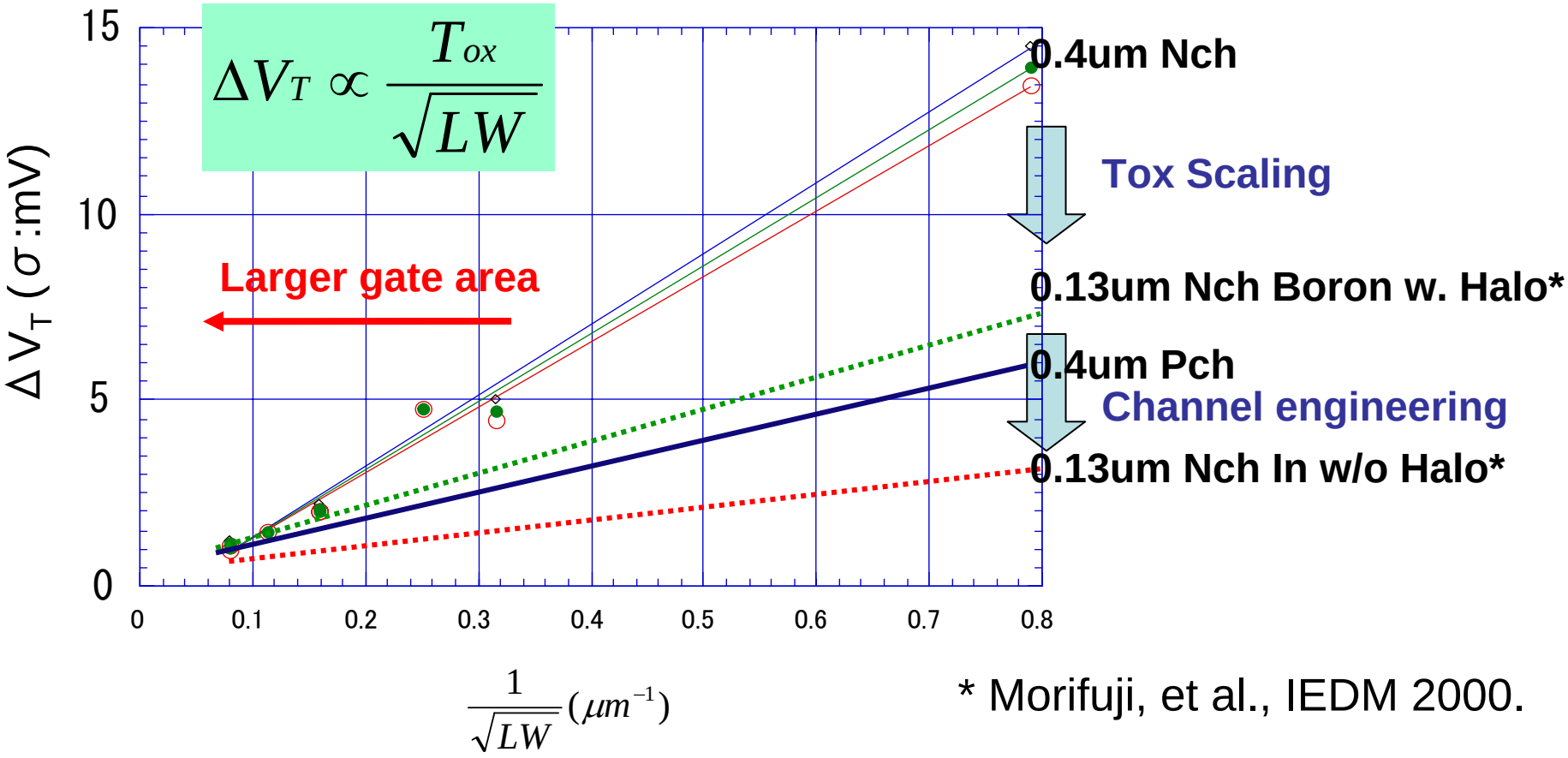
$$U_T \equiv \frac{kT}{q} \approx 26mV$$

$$gm_{CMOS} < \frac{1}{2}, \frac{1}{4} gm_{Bip}$$ （Same operating current）

$$Cin_{CMOS} < \frac{1}{2}, \frac{1}{4} Cin_{Bip}$$ （Same f_T）

V_T mismatch

V_T mismatch degrades accuracy; ADC, OP amp, and Mixer.
Larger gate area is needed for small V_T mismatch.
Scaling and proper channel structure improves mismatch.



* Morifuji, et al., IEDM 2000.

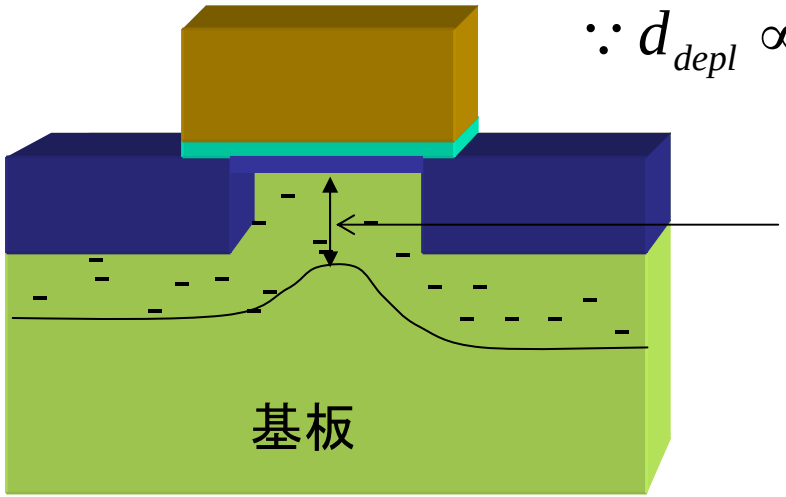
V_T mismatch: Fluctuation of doping

Courtesy of Prof. Taniguchi, Osaka Univ.

$$\Delta V_T = \frac{\Delta Q_{depl}}{C_{ox}} = A t_{ox} \frac{\sqrt{LW d_{depl} N_A}}{LW} = A' t_{ox} \frac{\sqrt[4]{N_A}}{\sqrt{LW}} \approx A_{VT} \frac{t_{ox}}{\sqrt{LW}}$$

$$\because d_{depl} \propto \frac{1}{\sqrt{N_A}}$$

$A_{VT} = 1V$



$$L = W = 0.25\mu m, \quad t_{ox} = 5nm$$

$$\Delta V_T = 20mV$$

T.Mizuno, J.Okamura and A.Toriumi, ”Experimental study o f threshold voltage fluctuation due to statistical variation of channel dopant number in MOSFETs,” IEEE Trans. On Electron Devices, ED-41, 2216 (1994)

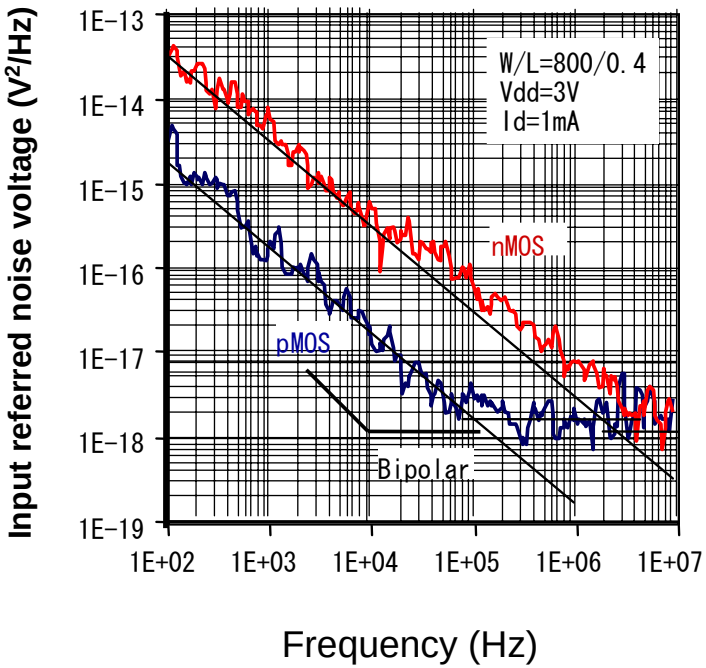
1/f noise

1/f noise of MOS is larger than that of bipolar.

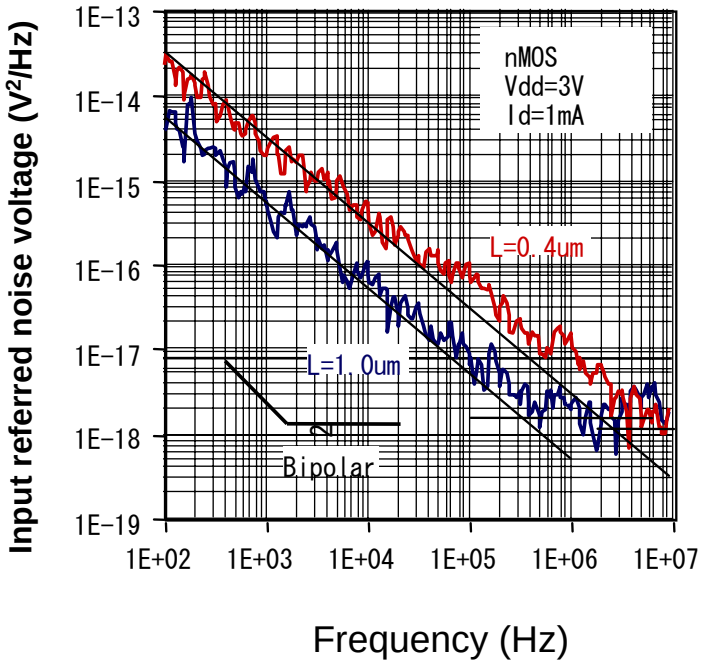
For the lower 1/f noise, the larger gate area is needed.

$$V_{nf}^2 = \frac{S_{vf}}{LW}\frac{\Delta f}{f}, \qquad S_{vf} \propto T_{ox}^2$$

Nch/Pch 0.4um

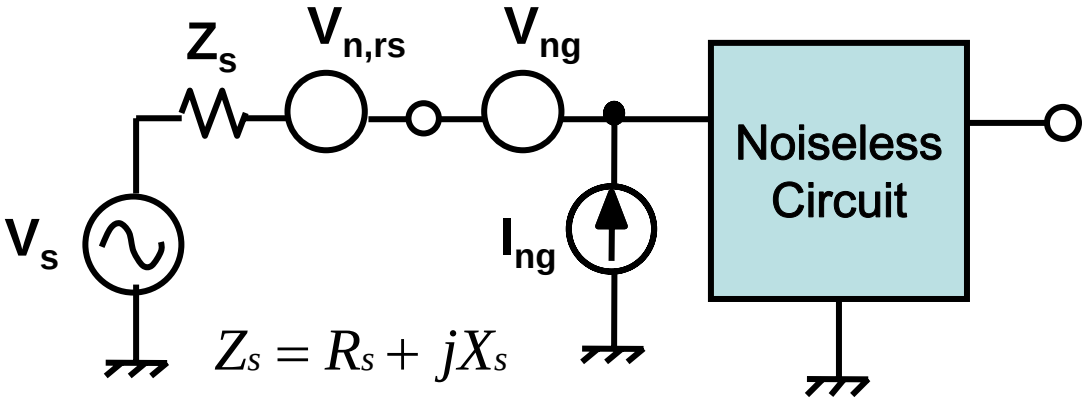


Nch 0.4um/1.0um



Noise figure: General

The lower R_{nv} and G_{ni} realizes the better for a lower noise figure.



$$Z_s = R_s + jX_s$$

$$\overline{V_{ng}{}^2} = 4kTR_{nv}, \quad \overline{I_{ng}{}^2} = 4kTG_{ni}$$

$$F = \frac{\overline{V_{n,rs}^2} + \overline{\left(V_{ng} + Z_s I_{ng}\right)^2}}{\overline{V_{n,rs}^2}} = 1 + \frac{R_{nv}}{R_s} + \frac{\left|Z_s\right|^2 G_{ni}}{R_s} \approx 1 + \frac{R_{nv}}{R_s} + R_s G_{ni}$$

$$R_{sopt} = \frac{\overline{V_{ng}}}{\overline{I_{ng}}} = \sqrt{\frac{R_{nv}}{G_{ni}}} \qquad F_{\min} \approx 1 + 2\sqrt{R_{nv} G_{ni}}$$

Noise figure: MOS transistor

$$F \approx 1 + \frac{R_{nv}}{R_s} + R_s G_{ni}$$

$$R_{nv} = R_g + R_{gs} \qquad R_g = R_{sr} \frac{W_{tot}}{L} \frac{1}{3N^2} \qquad R_{gs} \approx \frac{1}{5gm} \qquad G_{ni} \approx \frac{gm}{5} \left(\frac{\omega_0}{\omega_T} \right)^2$$

$$F \approx 1 + \frac{1}{Rs5gm} + Rs \frac{gm}{5} \left(\frac{\omega_0}{\omega_T} \right)^2 \qquad R_{sopt} \approx \frac{1}{gm} \left(\frac{\omega_T}{\omega_0} \right) = \frac{1}{C_{gs}\omega_0}$$

$$F_{\min} \approx 1 + 2\frac{\omega_0}{\omega_T}$$

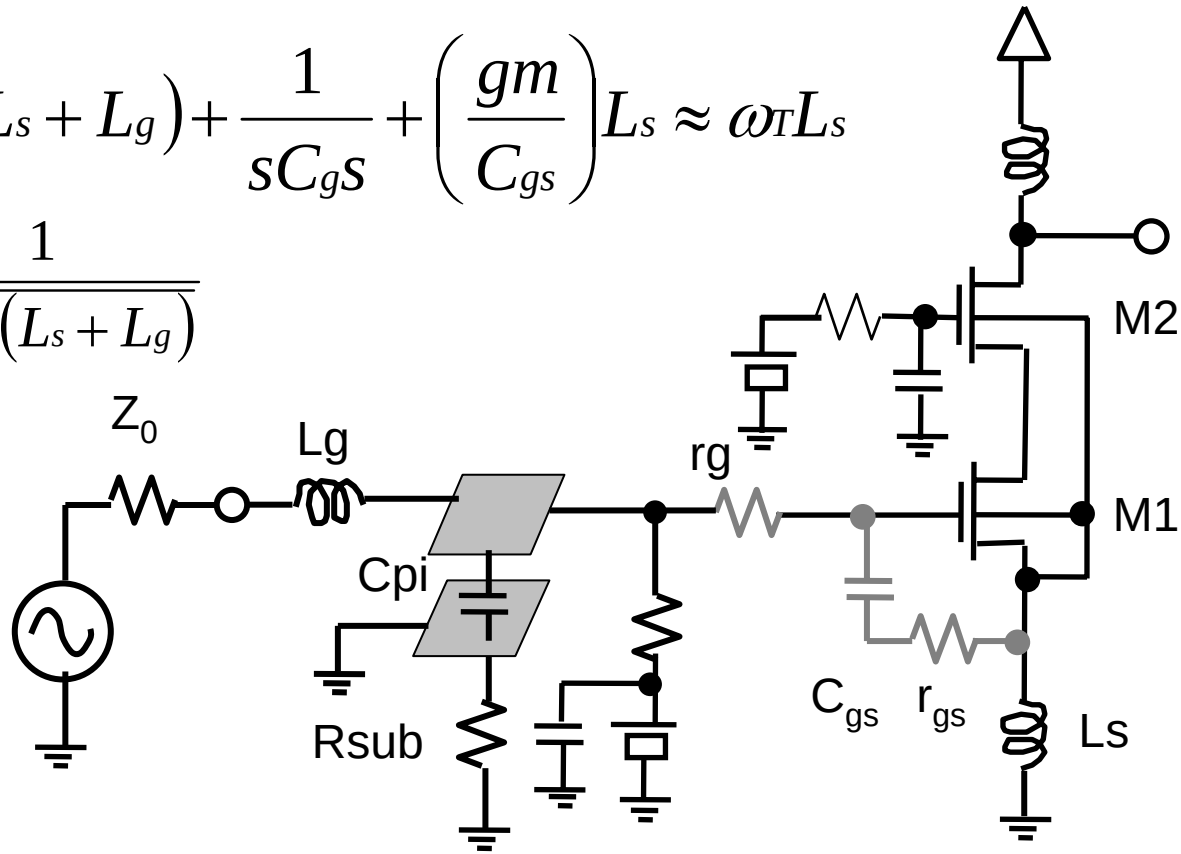
Low noise amplifier design

Narrowband LNA uses inductor degeneration for impedance matching.

Impedance matching

$$Z_{in} \approx s(L_s + L_g) + \frac{1}{sC_gs} + \left(\frac{gm}{C_{gs}}\right)L_s \approx \omega_T L_s$$

$$\omega_0 = \frac{1}{\sqrt{C_{gs}(L_s + L_g)}}$$



Low NF design

$$F \approx 1 + \frac{r_{gs} + r_g}{Z_0} + 4\gamma g_m Z_0 \left(\frac{\omega_0}{\omega_T}\right)^2 \approx 1 + \frac{r_{gs} + r_g}{Z_0}$$

$$r_{gs} \approx \frac{1}{5g_m}$$

Low noise figure

1) Lower the gate resistance

   Dived the gate or lower the gate sheet resistance

$$r_g = R_{sr}\frac{W_{tot}}{L}\frac{1}{3N^2}$$

$$r_g = R_{sr}\frac{W_{tot}}{L}$$

Rsr: Sheet resistance

N : The # of division

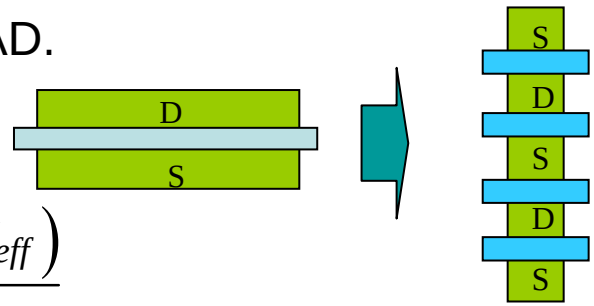
Divide the gate

2) Reduce substrate loss

   Reduce parasitic capacitance

   Use shield technique to the input bonding PAD.

   Use high resistive substrate, if possible.



3) Increase drain current

$$r_{gs} \approx \frac{1}{5g_m} \approx \frac{(V_{eff})}{10I_{ds}}$$

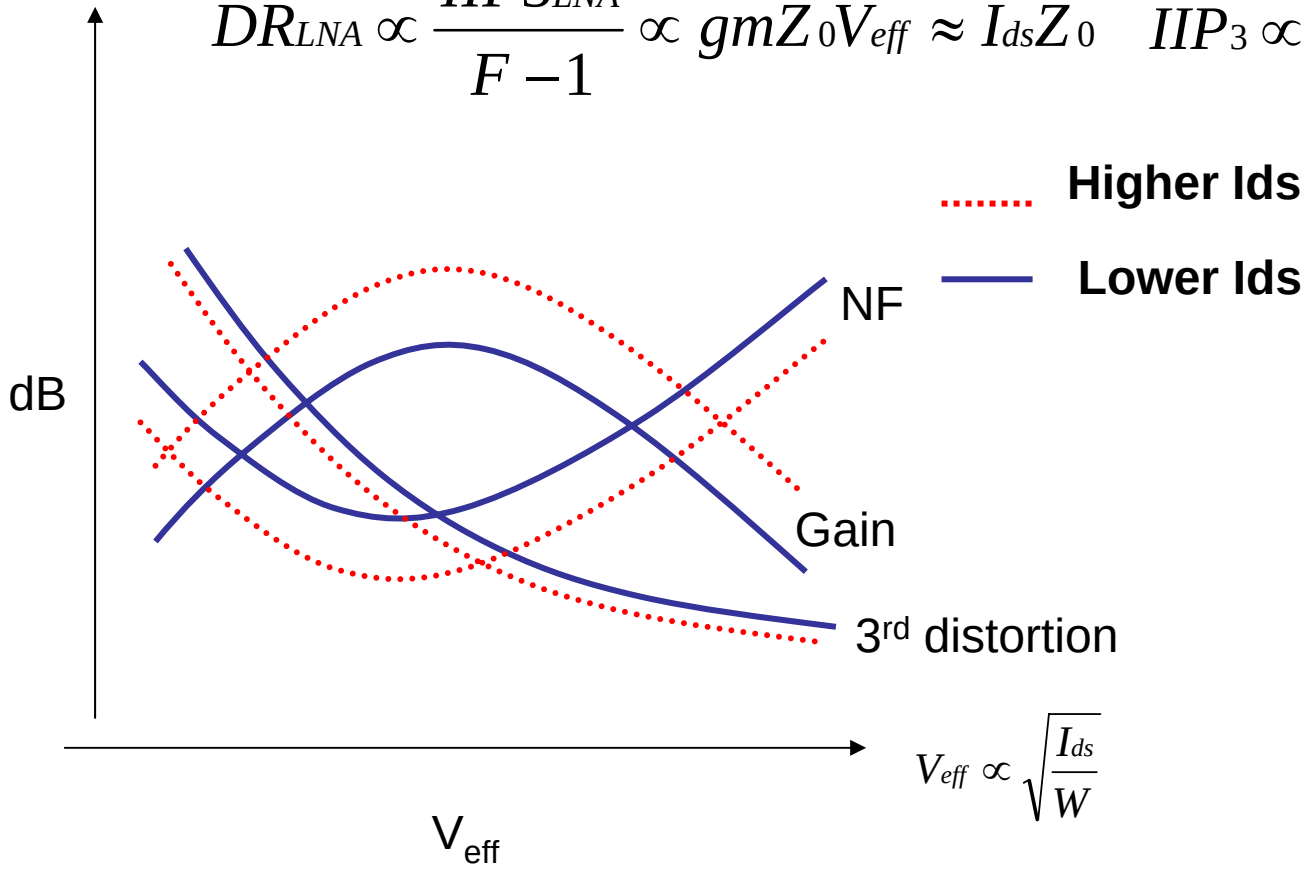
4) Increase Z_0, if possible.

I_{ds} and V_{eff} optimization

Adjust the I_{ds} and V_{eff} for optimization of gain, noise and distortion.

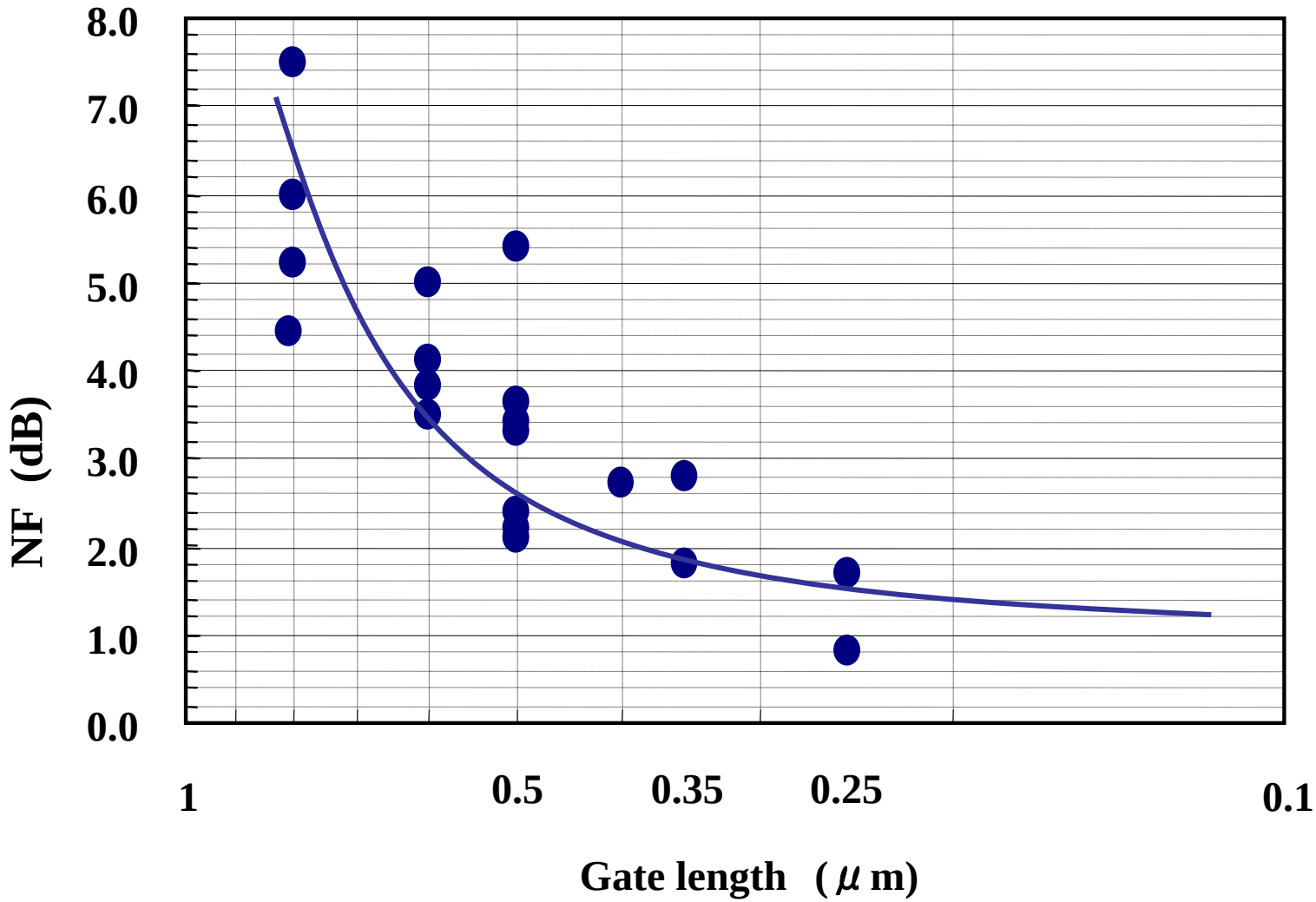
Dynamic range of LNA is proportional to Ids.

$$DR_{LNA} \propto \frac{IIP3_{LNA}}{F-1} \propto gmZ_0V_{eff} \approx I_{ds}Z_0 \quad IIP_3 \propto V_{eff}$$



NF progress in MOS LNA

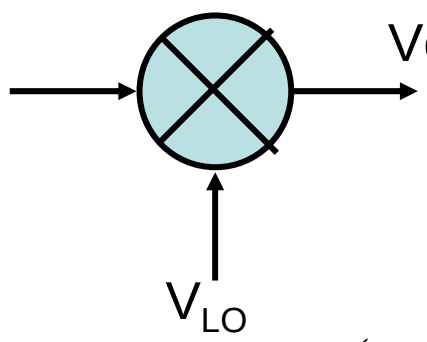
NF of MOS LNA is reaching 1dB.



Mixer

Mixer converts frequency, but image signal is converted to the same frequency.

$$V_s = A_s \cos(\omega_s t)$$



$$V_o = A_s \frac{2}{\pi} \cos((\omega_s \pm \omega_{LO})t)$$

If $V_{LO} >> 4V_{eff}$ (Full swing)

$$V_{LO} = A_{LO} \cos(\omega_{LO} t)$$

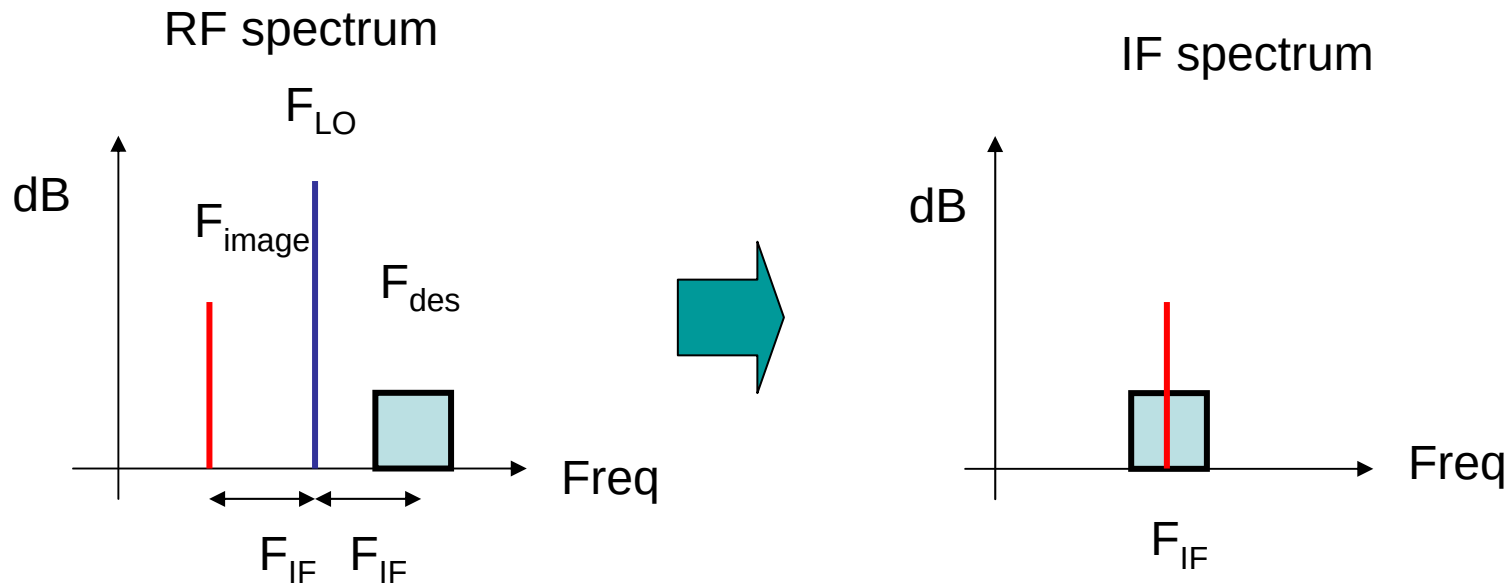
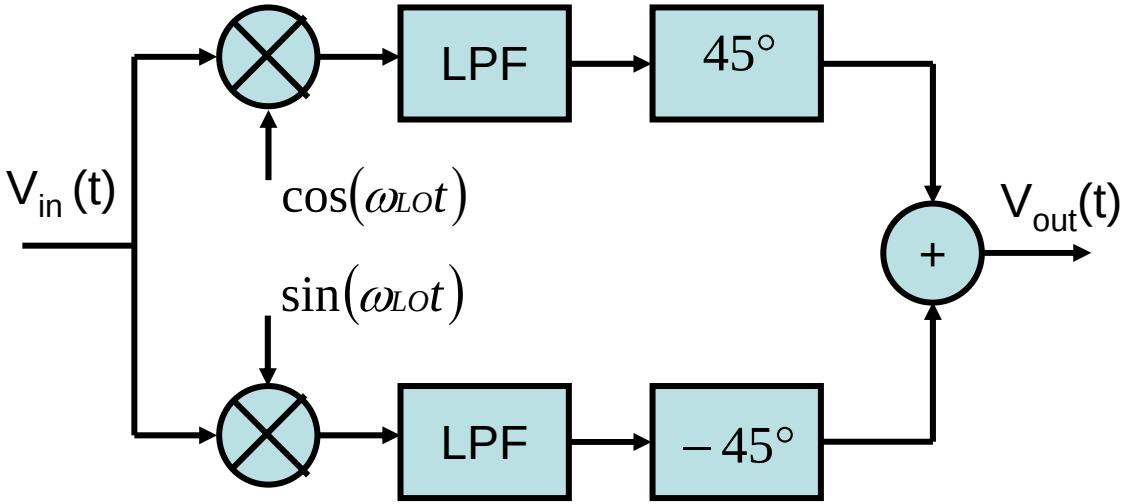


Image-reject mixers

The quadrature mixing realizes image-suppression.
Gain and phase matching is needed.



$$V_{in}(t) = A_{des}\cos(\omega_{des}t) + A_{im}\cos(\omega_{im}t)$$

$$V_{out}(t) = A_{des}A_c\cos(\omega_{IF}t) + A_{im}A_cI_R\cos(\omega_{IF}t)$$

A_c: Conversion gain, I_R: Image rejection

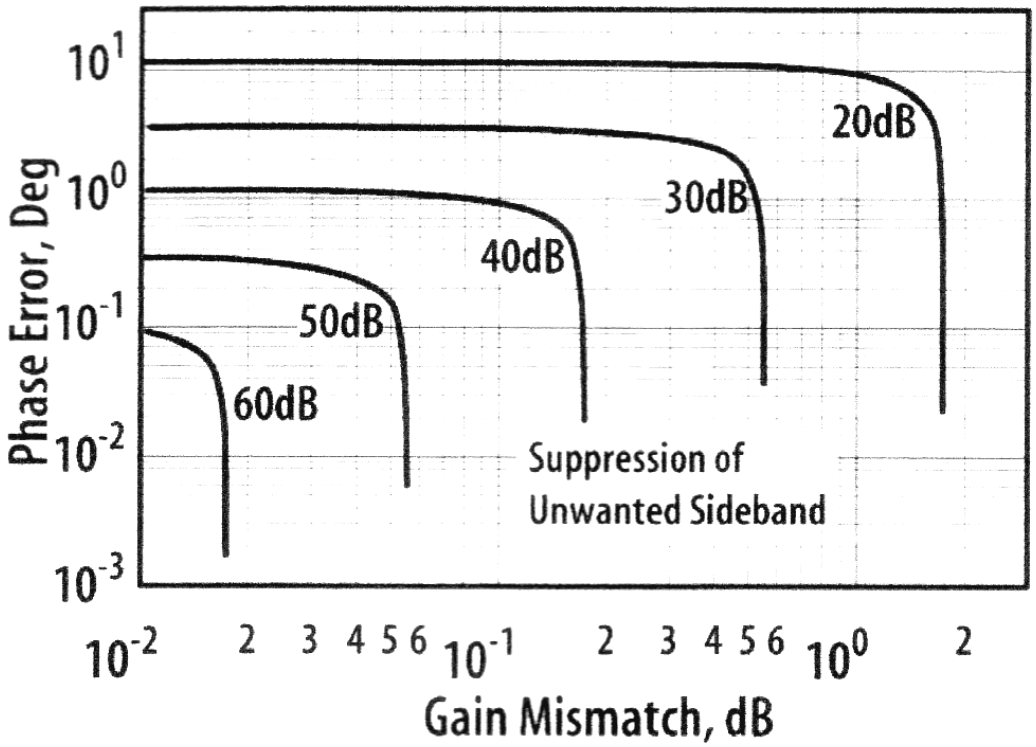
I_R=0 if I/Q phase difference is 90° and Channel conversion gains are equal.

Gain mismatch and phase error

$$\frac{P_{spur}}{P_{desired}} = \frac{1+\gamma^2-2\gamma\cos\phi}{1+\gamma^2+2\gamma\cos\phi}$$

γ : Gain ratio

ϕ :Phase error



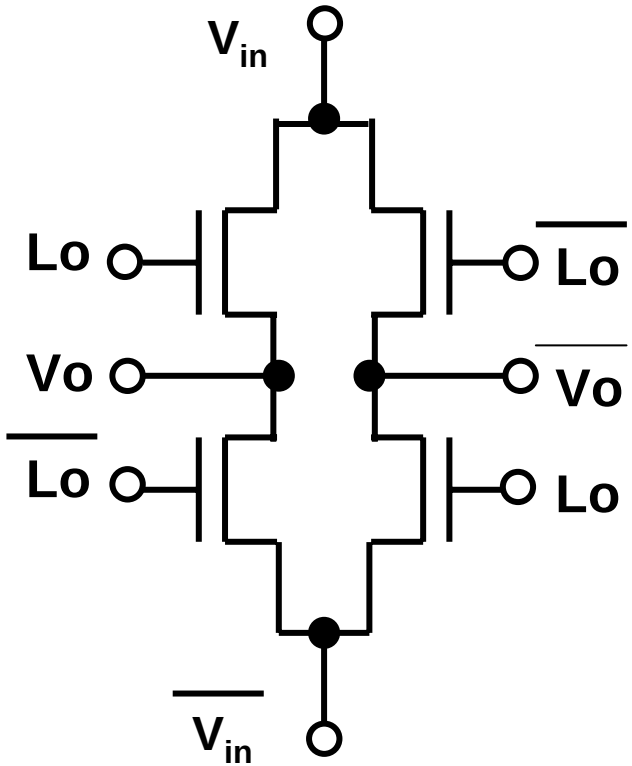
A. Rofougaran, et al., IEEE J.S.C. Vol.33, No.4, April 1998. PP. 515-534.

Passive FET mixer

MOS can realize a passive mixer easily.

Ultimately low power, but take care of isolation.

Passive FET mixer

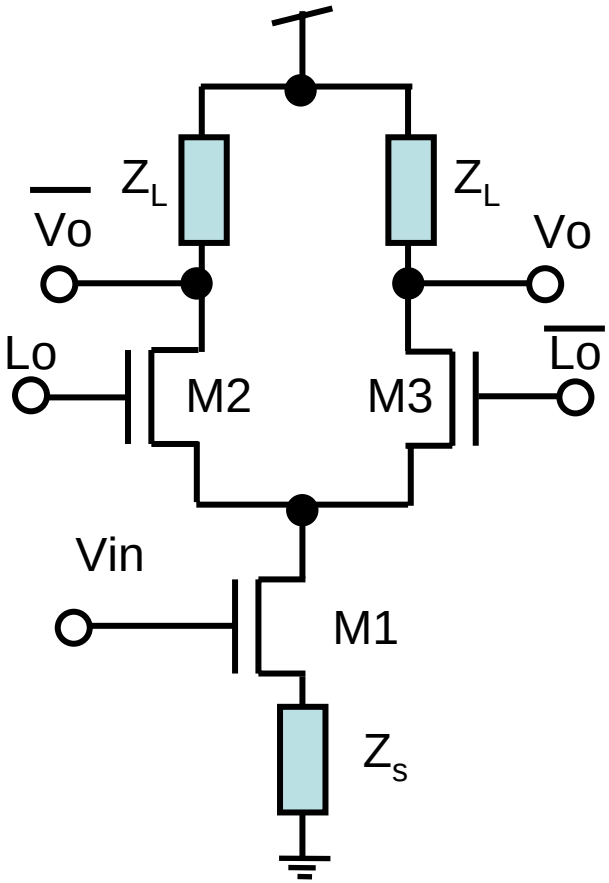


Low power
High linearity
No 1/F noise

No conversion gain
No isolation, Bi-directional

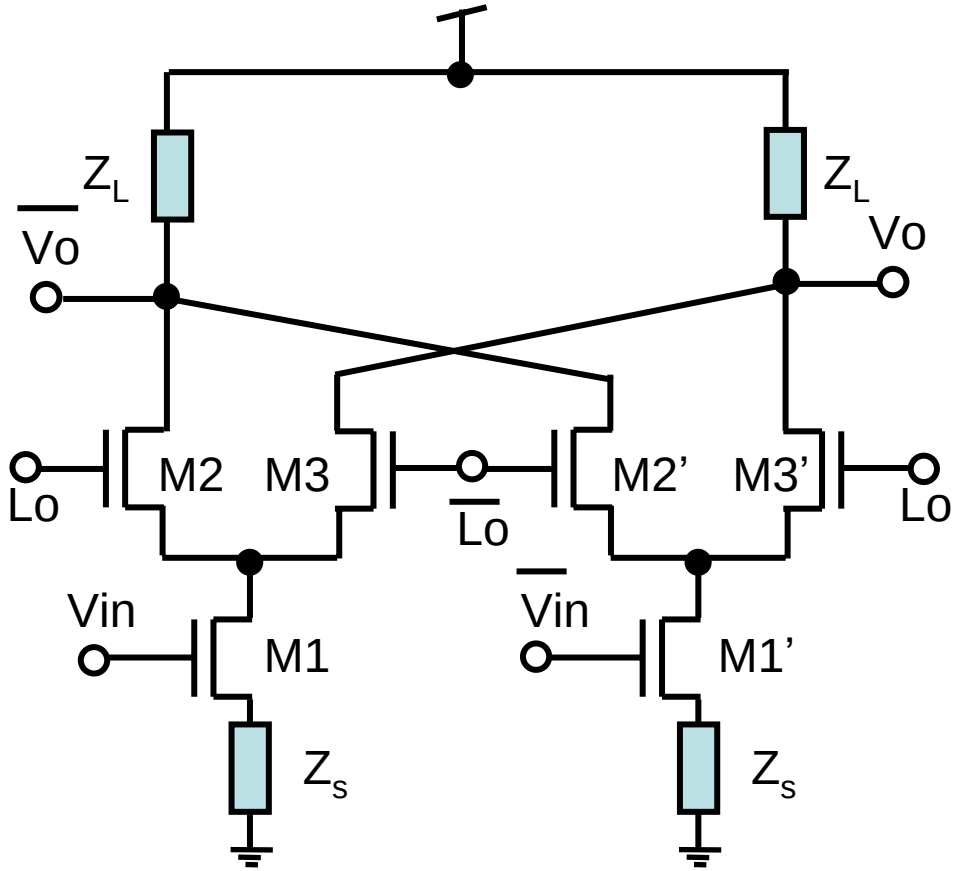
Active mixers

Single balanced mixer



Double balanced mixer

Very small direct feed through and even order distortion



Active mixer design

The larger Ids is needed for high dynamic range and shorter switching time for low 1/f noise.

Mixer gain $G_{mix} = \frac{2}{\pi} gm_1 Z_L, or = \frac{2}{\pi}\frac{Z_L}{Z_s}$ when Z_s is used

R_L: Resistive component in ZL

Thermal noise

$$v_{on}^2 = 8kTR_L\left(1+\frac{2\gamma IR_L}{\pi A_{LO}}+\gamma gm_1 R_L\right) \approx 8kTR_L^2\gamma gm_1$$

$$SSBv_{in}^2 = \frac{v_{on}^2}{\left(\frac{2}{\pi} gm_1 R_L\right)^2} \approx 2\pi^2 kT\frac{\gamma}{gm_1} \cong \pi^2 kT\gamma\frac{V_{eff}}{I_{ds}}$$

$$IIP3 \approx V_{eff}$$

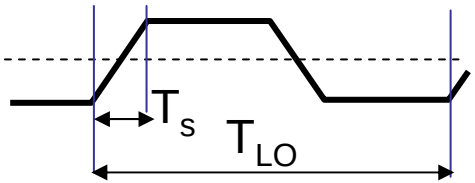
A larger dynamic range needs larger current

1/F noise

1) Switch transistor (M2, M3)

$$v_{n,o} = \frac{4Ts}{T_{LO}} v_{n,sw}$$

$$v_{n,sw}^2 \approx \frac{1}{WL} \propto \frac{1}{Cgs}$$



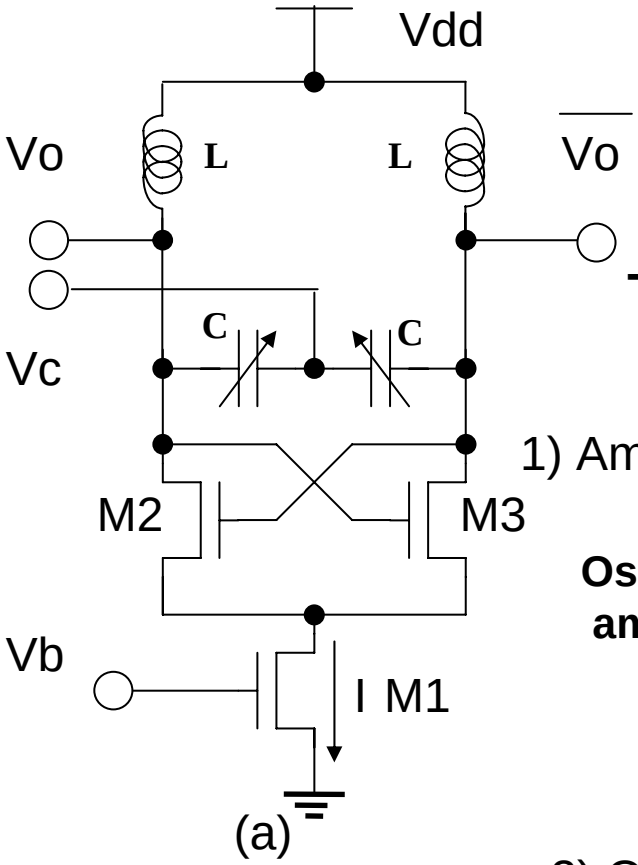
Phase modulation

Shorter switching time or larger T_s/T_{Lo} ratio

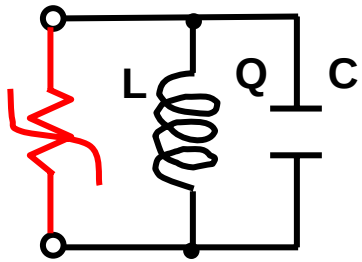
2) Load transistors Directly produces

Oscillator

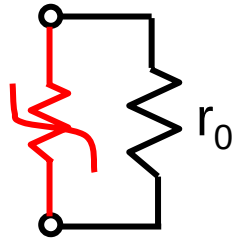
There is an optimum Ids for low phase noise.



-1/gm



-1/gm



$$r_o = Q\omega_0 L = \frac{Q}{\omega_0 C}$$

1) Amplitude condition

Oscillation amplitude $V_{osc} = \frac{4Ir_o}{\pi}$ **Headroom limit** 2Vdd

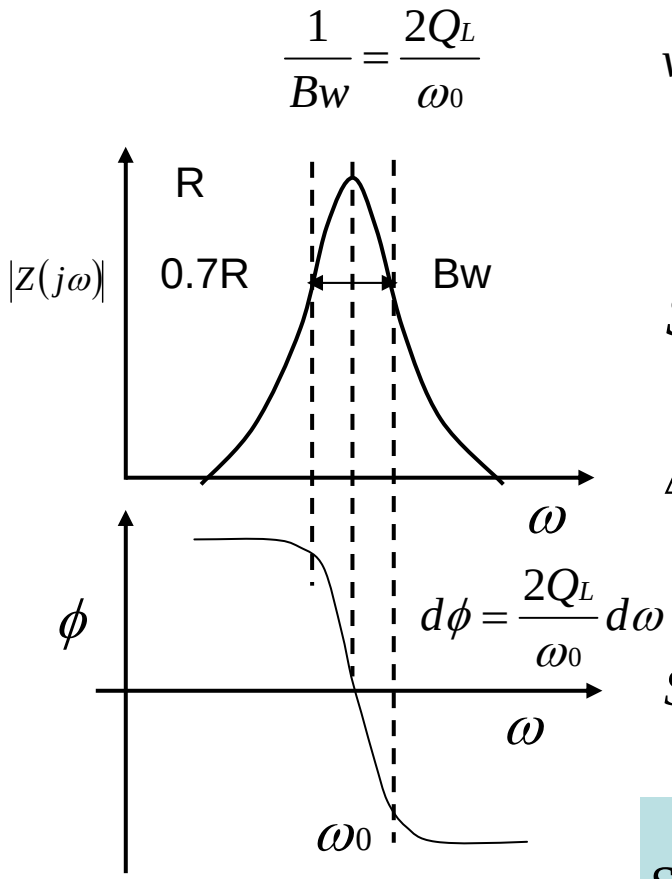
$$I_{opt} = \frac{\pi V_{dd}}{2r_o} = \frac{\pi V_{dd}\omega_o C}{Q}$$

2) Oscillation condition

$$gm_{2,3} > \frac{2}{r_o}, \qquad I > \frac{\omega_o C V_{eff,2,3}}{Q}$$

Phase noise of oscillator

Phase-frequency relation and resonator characteristics determine phase noise.



$$v(t) = A\cos\left[\omega_0 t + \phi(t)\right]$$

$$\omega_m = \frac{d\phi}{dt} = j\omega\phi$$

$$S_\omega(\omega_m) = \omega_m^2 S_\phi(\omega_m)$$

$$\Delta\theta = \frac{\omega_m}{Bw} = \frac{2Q_L}{\omega_0}\omega_m$$

$$S_\omega(\omega_m) = \left(\frac{\omega_0}{2Q_L}\right)^2 S_{\Delta\theta}(\omega_m) \qquad \omega_m < Bw$$

$$S_\phi(\omega_m) = \left(\frac{\omega_0}{2Q_L}\right)^2 \frac{1}{\omega_m^2} S_{\Delta\theta}(\omega_m) = \left(\frac{\omega_0}{2Q_L\omega_m}\right)^2 S_{\Delta\theta}(\omega_m)$$

ω_m : Offset angular frequency

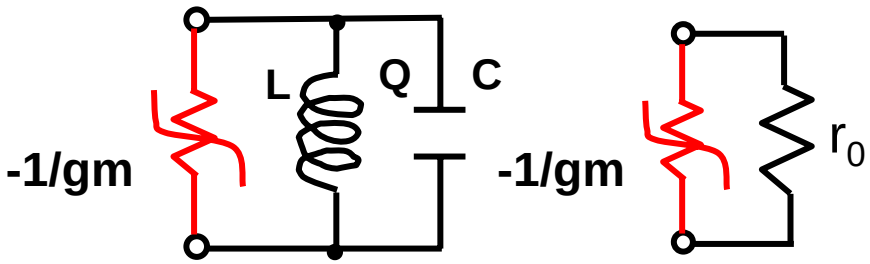
$S_\omega(\omega_m)$:Noise spectrum density on offset angular frequency

$S_\phi(\omega_m)$:Noise spectrum density on phase

$\Delta\theta$ Phase error between in and out

$S_{\Delta\theta}(\omega_m)$:Noise spectrum density on phase error

Phase noise of oscillator



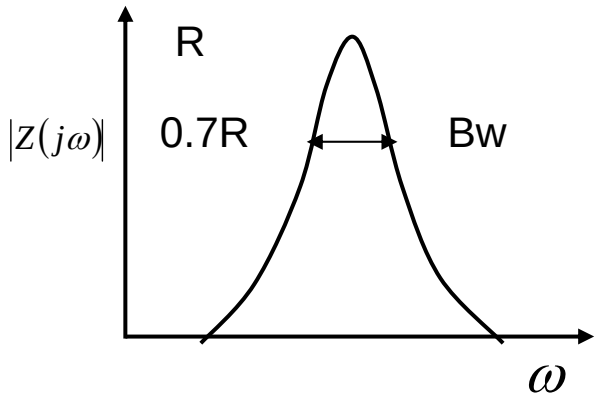
$$Z(\omega_0+\omega_m)\approx j\frac{\omega_0 L}{2\frac{\omega_m}{\omega_0}} \qquad \omega_m << \omega_0$$

(Filter action)

$$Q=\frac{r_0}{\omega_0 L}$$

$$\left|Z(\omega_0+\omega_m)\right|\approx\frac{r_0\omega_0}{2Q\omega_m}$$

$$\frac{1}{Bw}=\frac{2Q}{\omega_0}$$



$$\frac{\overline{v_n^2}}{\Delta f}=\frac{\overline{i_n^2}}{\Delta f}\cdot|Z|^2=4kTr_o\left(\frac{\omega_0}{2Q\omega_m}\right)^2$$

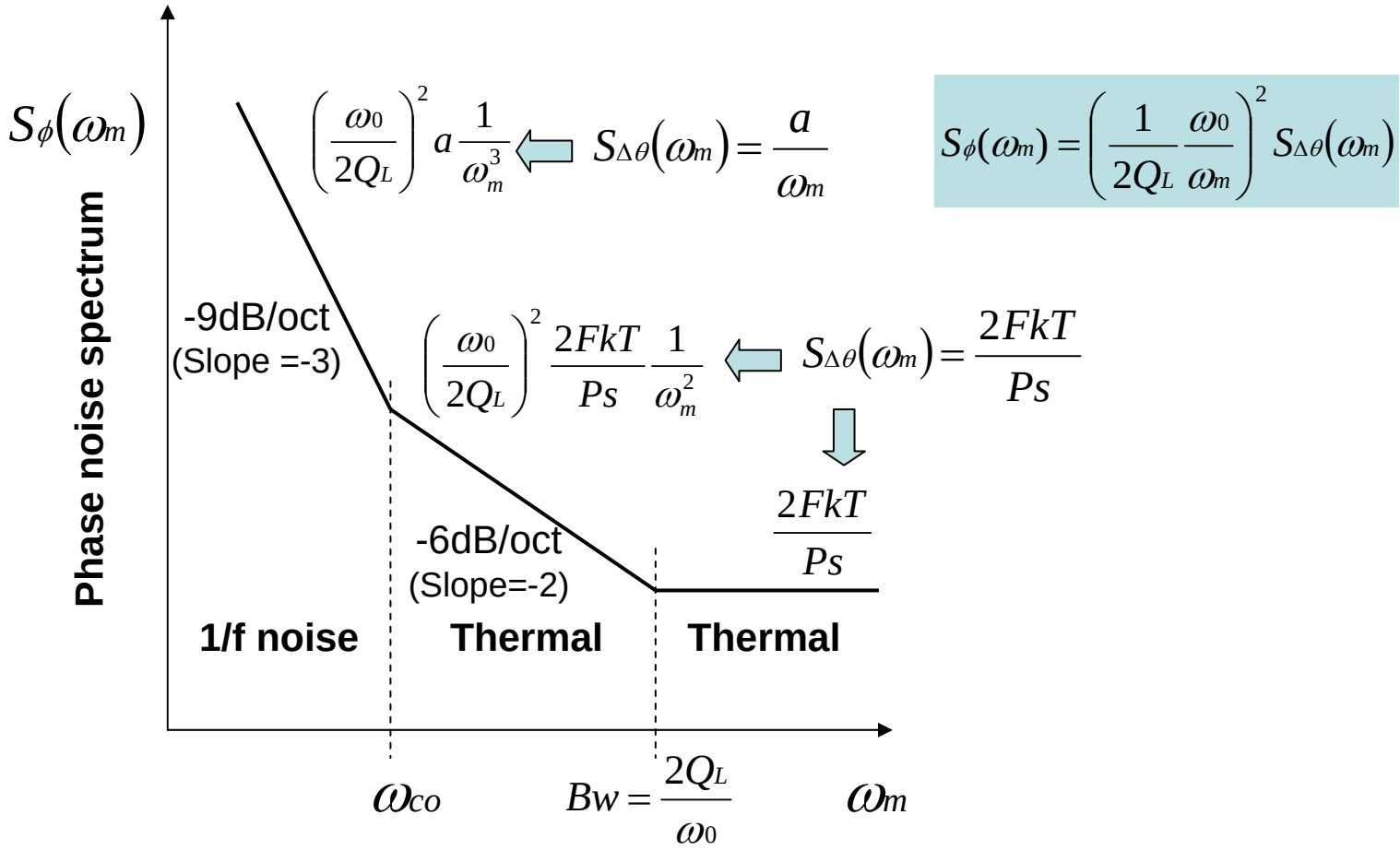
Noise spectrum density

$$L\{\omega_m\}=10\log\left[\frac{2kT}{P_{sig}}\cdot\left(\frac{\omega_0}{2Q\omega_m}\right)^2\right]$$

Phase noise

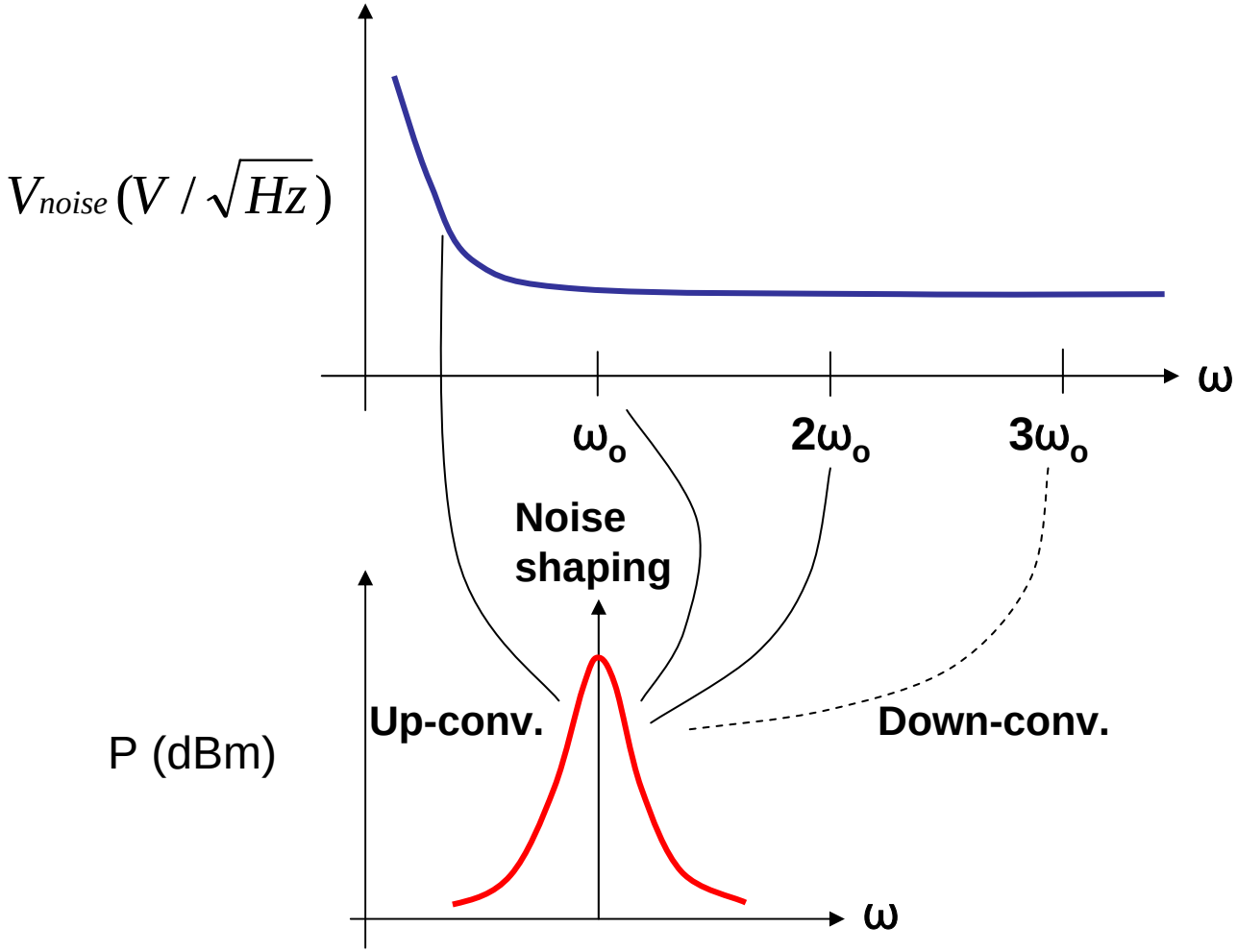
Frequency characteristics of Phase noise in oscillator

1/f noise and thermal noise is converted to $1/f^3$ and$1/f^2$, respectively.



Up and down converted noise

Noises around N*fo are up and down converted to fo.



FoM and minimum phase noise

FoM is basically proportional to Q².

$$FoM = \left(\frac{f_0}{f_m}\right)^2 \frac{1}{L(f_m)V_{dd}I}$$

Fm: Offset frequency
L(fm): Phase noise at offset freq.

$$L(f_m) = \frac{1}{2}\cdot\frac{1}{Q^2}\cdot\left(\frac{f_o}{f_m}\right)^2\cdot\frac{FkT}{P_{RF}} = \frac{1}{2}\cdot\frac{1}{Q^2}\cdot\left(\frac{f_o}{f_m}\right)^2\cdot\frac{FkT}{\left(\frac{V_o^2}{2ro}\right)}$$

F: Noise factor

$$F = 2 + \frac{8\gamma r_o I}{\pi V_o} + \gamma\frac{8}{9}r_o\cdot g_{m1} \qquad I_{opt} = \frac{\pi V_{dd}}{2r_o} = \frac{\pi V_{dd}\omega_o C}{Q} = \frac{\pi V_{dd}}{2Q\omega_o L_{ind}}$$

$$FoM = \frac{4}{\pi}\frac{Q^2}{kT}\frac{1}{2+4\gamma+\frac{32}{9}\gamma\pi\frac{V_{dd}}{V_{eff,1}}} \propto Q^2$$

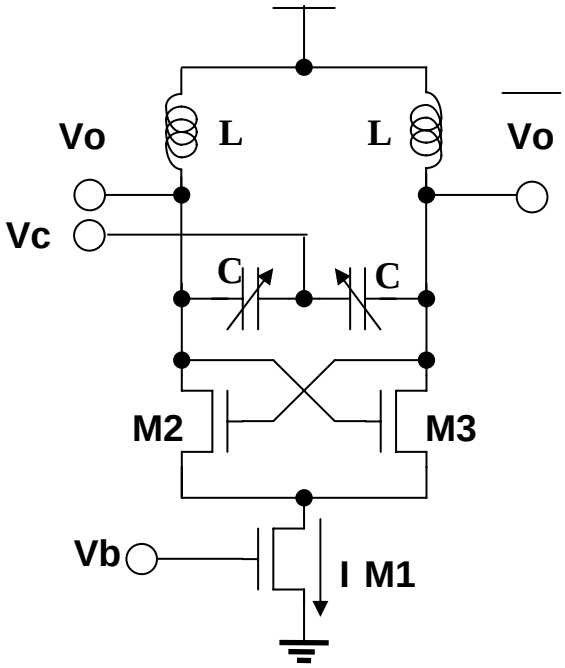
at I_{opt}

Oscillator design

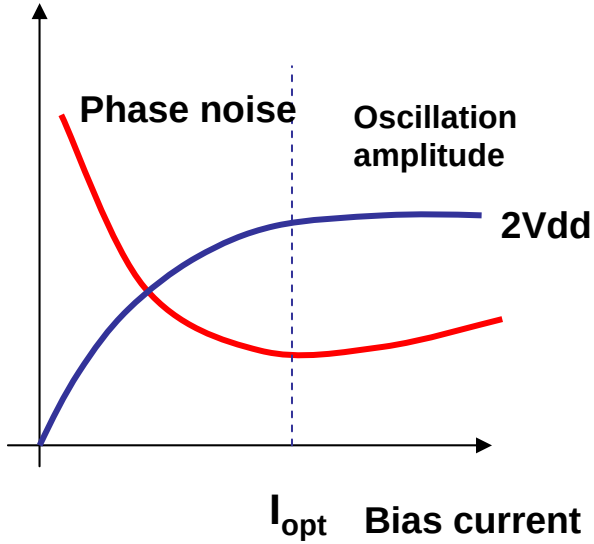
Careful optimization reduces the oscillator phase noise.

$$L_{\min}(f_m) = kT \cdot \frac{\gamma}{V_{dd}} \cdot \frac{\omega_o L_{ind}}{2Q} \cdot \left(\frac{1}{V_{dd}} + \frac{2}{V_{eff,1}} \right) \left(\frac{f_o}{f_m} \right)^2$$

$$L_{\min}(f_m) = kT \cdot \frac{\gamma}{2I_{opt}} \cdot \frac{1}{2Q^2} \cdot \left(\frac{1}{V_{dd}} + \frac{2}{V_{eff,1}} \right) \left(\frac{f_o}{f_m} \right)^2$$



$$I_{opt} = \frac{\pi V_{dd}}{2r_o} = \frac{\pi V_{dd} \omega_o C}{Q} = \frac{\pi V_{dd}}{2Q\omega_o L_{ind}}$$



Larger V_{dd}
Large V_{eff1}, but take care of V_o reduction
Large L_1, W_1 to reduce 1/f noise
Enough W/L for M2, M3
Higher Q
Larger QL_{ind} for Lower I_{opt}

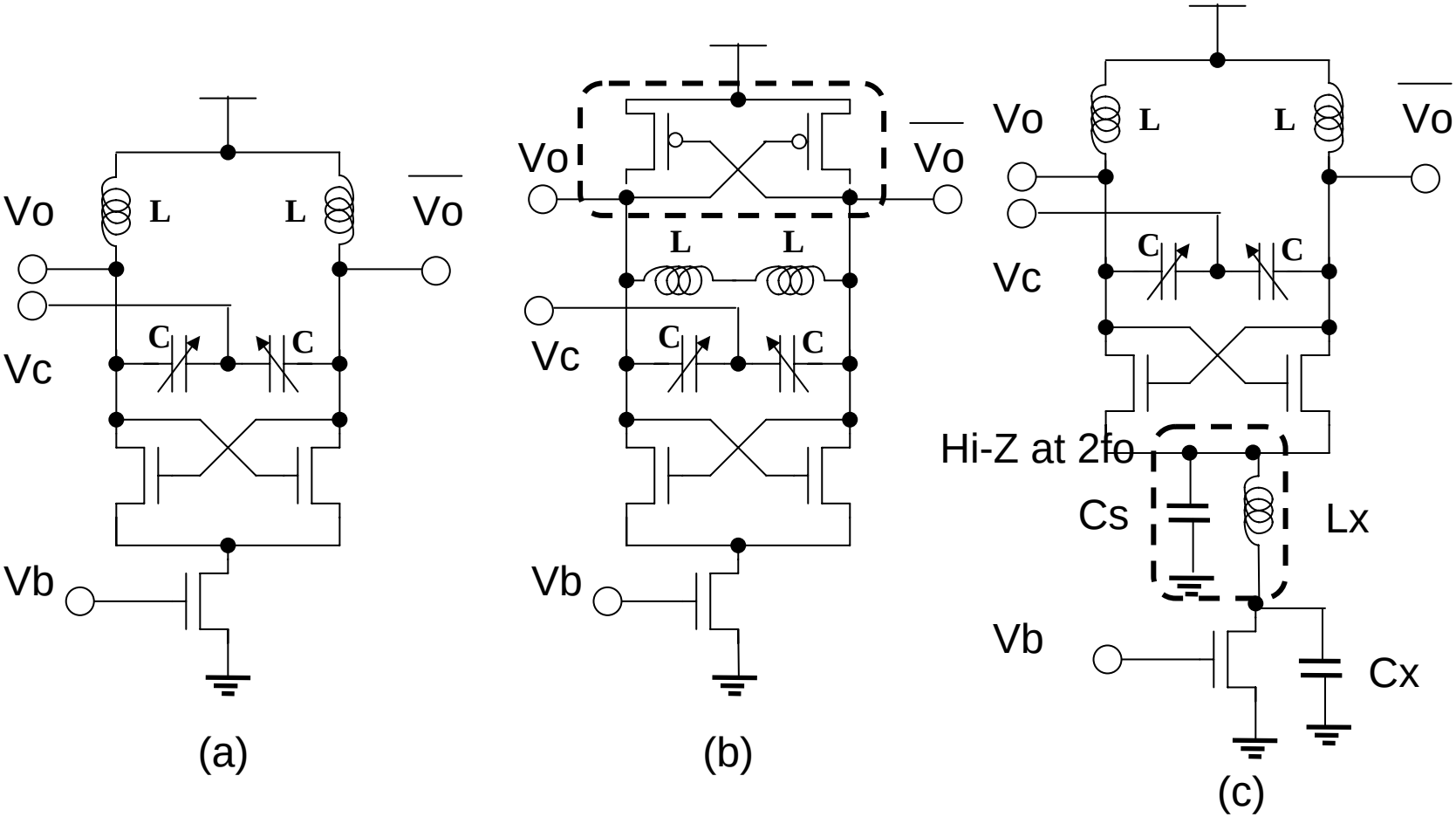
CMOS oscillator circuits

E. Hegazi, ISSCC 2001

Basic

Low power (gm is higher)

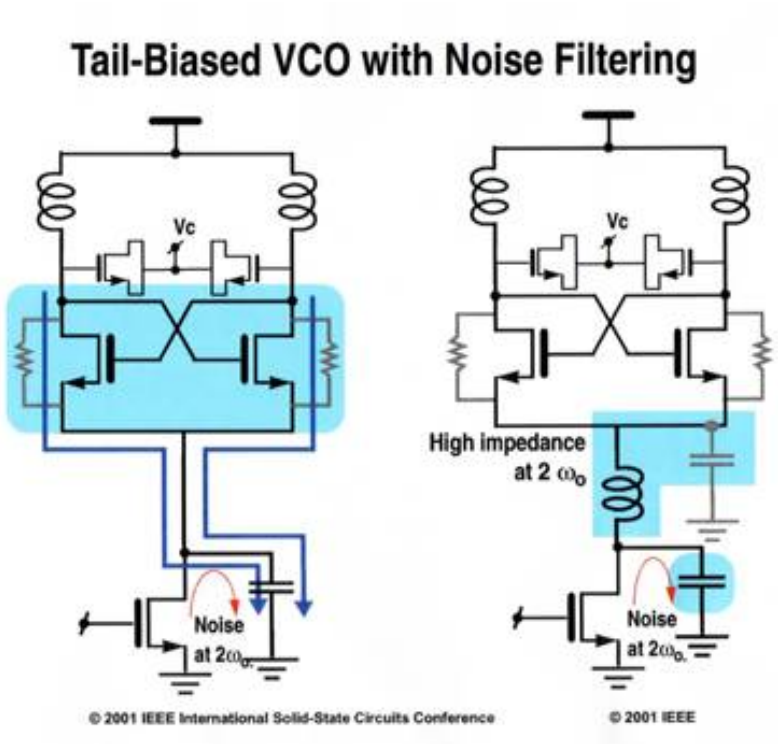
Low noise by filtering

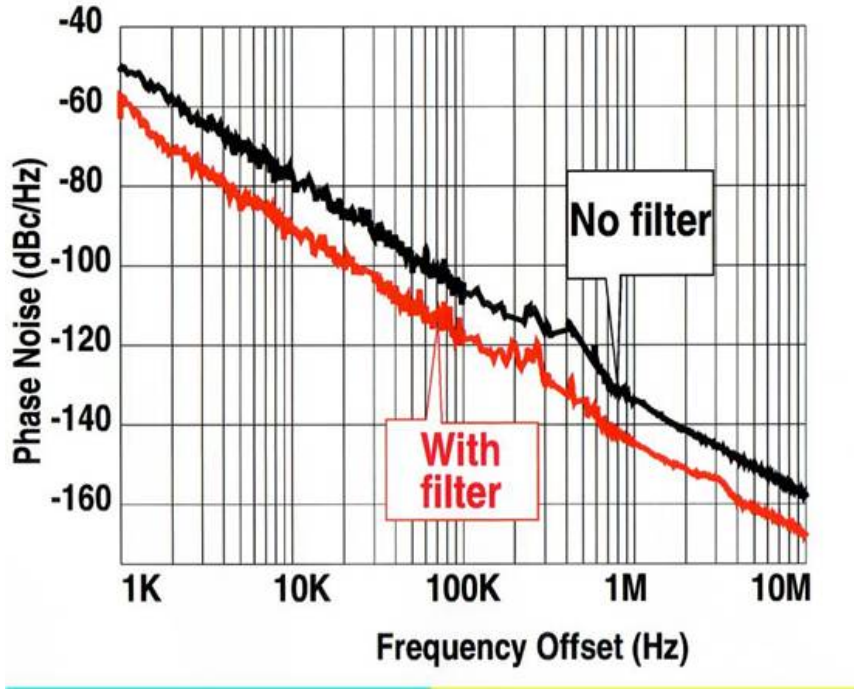


Filtering of $2f_o$ component in OSC.

Noise filtering of $2f_o$ component reduces the OSC phase noise to -10dB.

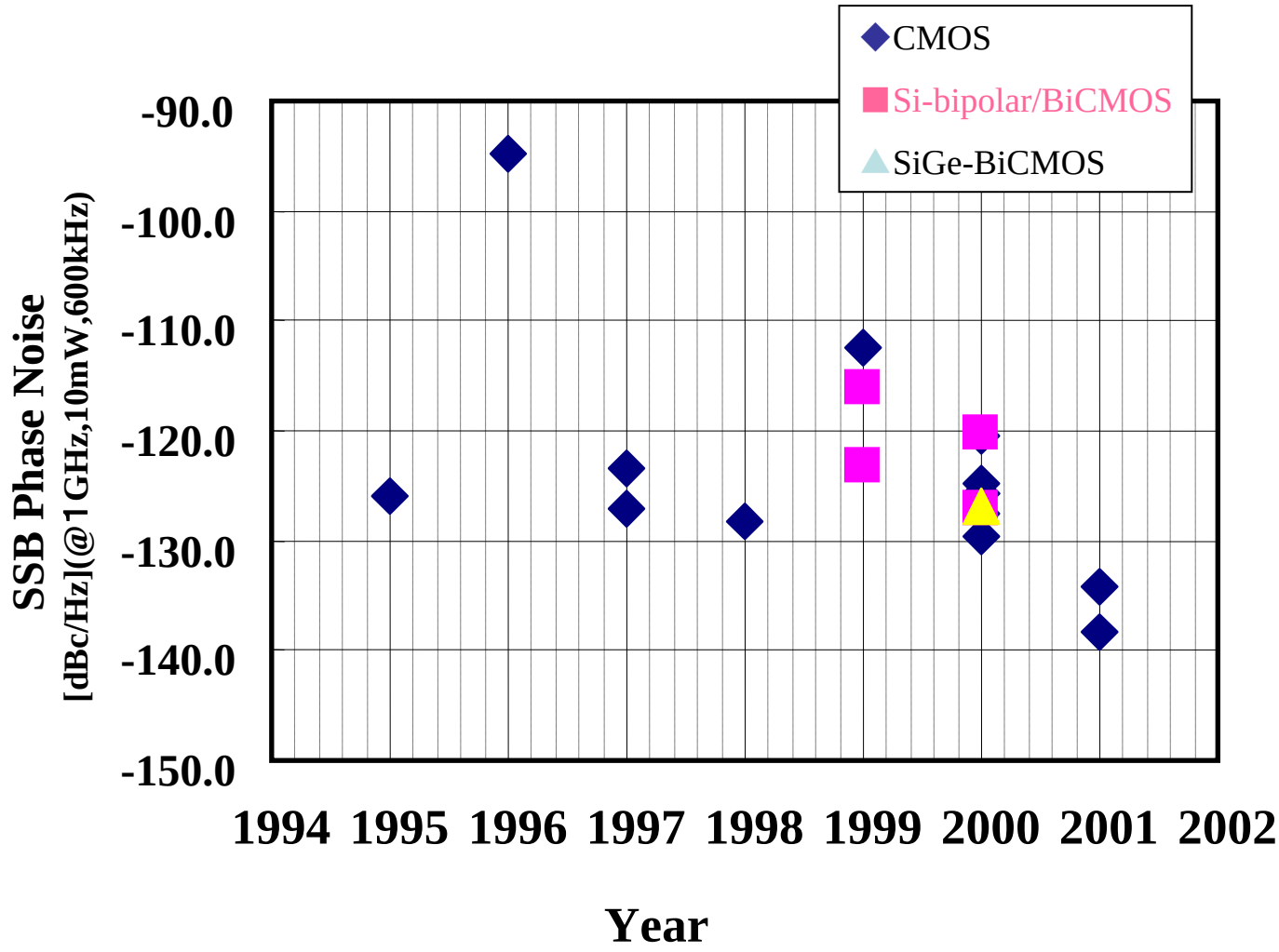
E. Hegazi, ISSCC 2001





Oscillator phase noise progress

Phase noise in CMOS oscillator becomes lower than that of bipolar.



Acknowledgment and references

- **Acknowledgment**

I would like to thank Prof. Asad Abidi in UCLA for his advices.

- **References**

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  - Thomas. H. Lee, “The design of CMOS RF ICs,” Cambridge University Press, Jan. 1998.
  - Bezad, Razavi, “RF micro-electronics,” Prentice Hall, Nov. 1999.
  - Domine Leenaerts, Johan van der Tang, and Ciero Vaucher, “Circuit Design for RF Transceivers,” Kluwer Academic Publishers, 2001.
  - Charles Chien, “Digital Radio Systems on A chip,” Kluwer Academic Publishers, 2001.
  - E. Hegazi, et. Al., ”A Filtering Technique to Lower Oscillator Phase Noise,” ISSCC 2001, 23.4, Feb. 2001.

An In-Line RF Wattmeter

Considerable attention was devoted to the resistance-type SWR bridge in the preceding section because it is the simplest type that is capable of adequate accuracy in measuring standing-wave ratio. Its disadvantage is that it must be operated at a very low power level, and thus is not suitable for continuous monitoring of the SWR in actual transmission. To do this the instrument must be capable of carrying the entire power output of the transmitter, and should do it with negligible loss. An RF wattmeter meets this requirement.

It is neither costly nor difficult to build an RF wattmeter. And, if the instrument is equipped with a few additional components, it can be switched to read reflected power as well as forward power. With this feature the instrument can be used as an SWR meter for antenna matching and Transmatch adjustments. The wattmeter shown in Figs 9 through 12 meets these requirements. The instrument uses a directional type of coupler for sampling the energy on the transmission line. The indicator sensitivity of this instrument is not related to frequency, as is the case with some types of directional couplers. This unit may be calibrated for power levels as low as 1 watt, full scale, in any part of the HF spectrum. With suitable calibration, it has good accuracy over the 3-30 MHz range. It is built in two parts, an RF head for inserting in the coaxial transmission line, and a control-meter box which can be placed in any location where it can be operated conveniently. Only direct current flows in the cable connecting the two pieces.

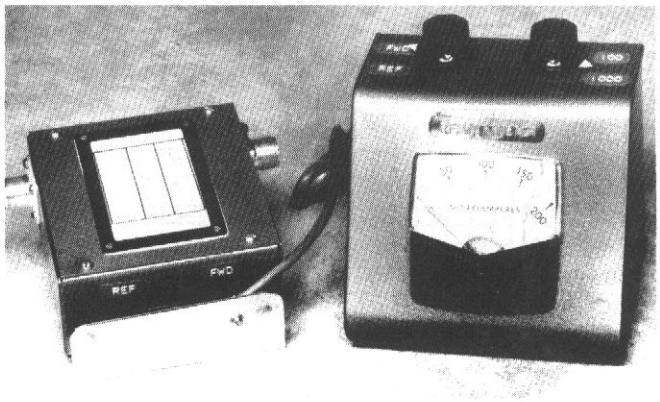


Fig 9—The RF wattmeter consists of two parts, the RF head (left), and the control-meter box (right). The paper scale affixed to the RF head contains the calibration information which appears in Fig 10.

Design Philosophy

See the circuit of Fig 10. The transmission line center conductor, W1, passes through the center of a toroid core and becomes the primary of T1. The multiturn winding on the core functions as the transformer secondary. Current flowing on W1 induces a voltage in the secondary which causes a current to flow through resistors R1 and R2. The voltage drops across these resistors are equal in amplitude, but 180° out of phase with respect to common or ground. They are thus, for practical purposes, respectively in and out of phase with the line current. Capacitive voltage dividers, C1-C3 and C2-C4, are connected across the line to obtain equal-amplitude voltages *in phase* with the line voltage, the division ratio being adjusted so that these voltages match the voltage drops across R1 and R2 in amplitude. (As the current/voltage ratio in the line depends on the load, this can be done only for a particular value of load impedance. Load values chosen for this standardization are pure resistances that match the characteristic impedance of the transmission line with which the bridge is to be used, 52 or 75 ohms usually.) Under these conditions, the voltages rectified by D1 and D2 represent, in the one case, the vector *sum* of the voltages caused by the line current and voltage, and in the other, the vector *difference*. With respect to the resistance for which the circuit has been set up, the sum is proportional to the forward component of a traveling wave such as occurs on a transmission line, and the difference is proportional to the reflected component.

Component Selection

R1 and R2 should be selected for the best null reading when adjusting the bridge into a resistive 52- or 75-ohm load. Normally, the value will be somewhere between 10 and 47 ohms. The 10-ohm value worked well with the instruments shown here. Half-watt composition resistors are suitable to 30 MHz. R1 and R2 should be as closely matched in resistance as possible. Their exact value is not critical, so an ohmmeter may be used to match them.

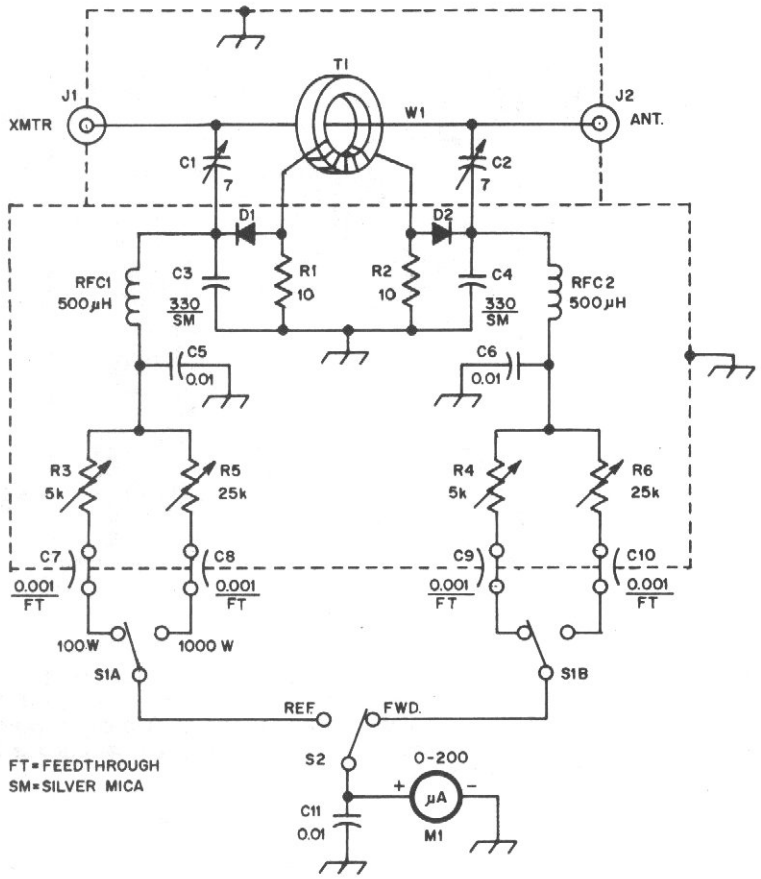
Ideally, C3 and C4 should be matched in value. Silver-mica capacitors are usually close enough in tolerance that special selection is not required, providing there is enough leeway in the ranges of C1 and C2 to compensate for any difference in the values of C3 and C4.

Diodes D1 and D2 should also be matched for best results. An ohmmeter can be used to select a pair of diodes having forward dc resistances within a few ohms of being the same. Similarly, the back resistances of the diodes can be matched. The matched diodes will help to assure equal meter readings when the bridge is reversed. (The bridge should be perfectly bilateral in its performance characteristics.) Germanium diodes should be used to avoid misleading results when low values of reflected power are present during antenna adjustments. The SWR can appear to be perfect when actually it isn't. The germanium diodes conduct at approximately 0.3 volt, making them more suitable for low-power readings than silicon diodes (conduction at 0.7 V).

Any meter having a full-scale reading between 50 μA and 1 mA can be used at M1. The more sensitive the meter, the more difficult it will be to get an absolute reflected-power reading of zero. Some residual current will flow in the bridge circuit no matter how carefully the circuit is balanced, and a sensitive instrument will indicate this current flow. Also, the more sensitive the meter, the larger will have to be the calibrating resistances, R3 through R6, to provide high-power readings. A 0-200 μA meter represents a good compromise for power ranges between 100 and 2000 watts.

Construction

It is important that the layout of any RF bridge be as symmetrical as possible if good balance is to be had. The circuit-board layout of Fig 12 meets the requirement for this instrument. The input and output ports of the equipment should be isolated from the remainder of the circuit so that only the sampling circuits feed voltage to the bridge. A shield across the end of the box which contains the input and output jacks and W1 is necessary. If stray RF gets into the bridge



WATTS	M1	WATTS
100	200	1000
90	180	900
80	170	800
70	155	700
60	145	600
50	125	500
40	105	400
30	85	300
20	65	200
10	40	100
5	20	50

Fig 10—Schematic diagram of the RF wattmeter. A calibration scale for M1 is shown also. Fixed-value resistors are ½-watt composition. Fixed-value capacitors are disc ceramic unless otherwise noted. Decimal-value capacitances are in microfarads. Others are picofarads. Resistance is in ohms; k = 1000.

C1, C2—1.3- to 6.7-pF miniature trimmer (E. F. Johnson 189-502-4, available from Newark Electronics, Chicago, Illinois)

C3-C11, incl—Numbered for circuit-board identification.

D1, D2—Matched small-signal germanium diodes, 1N34A, etc (see text).

J1, J2—Chassis-mount coax connector of builder's choice. Type SO-239 used here.

M1—0-200 μA meter (Triplett type 330-M used here).

R1, R2—Matched 10-Ω resistors (see text).

R3, R4—5-kΩ printed-circuit carbon control (IRC R502-B).

R5, R6—25-kΩ printed-circuit carbon control (IRC R242-B).

RFC1, RFC2—500-μH RF choke (Millen 34300-500 or similar).

S1—DPDT single-section phenolic wafer switch (Mallory 3222J).

S2—SPDT phenolic wafer switch (Centralab 1460).

T1—Toroidal transformer; 35 turns of no. 26 enam wire to cover entire core of Amidon T-68-2 toroid (available from Amidon Assoc or Radiokit).

W1—Numbered for text discussion.

circuit, it will be impossible to obtain a complete zero reflected-power reading on M1 even though a 1:1 SWR exists.

All of the RF head components except J1, J2 and the feedthrough capacitors are assembled on the board. The board is held in place by means of a homemade aluminum L bracket at the end nearest T1. The circuit board end nearest the feedthrough capacitors is secured with a single no. 6 spade bolt. Its hex nut is outside the box, and is used to secure a solder lug which serves as a connection point for the ground braid in the cable which joins the control box to the RF head.

T1 fits into a cutout area of the circuit board. A 1-inch long piece of RG-8 coax is stripped of its vinyl jacket and shield braid, and is snug-fit into the center hole of T1. The inner conductor is soldered to the circuit board to complete the W1 connection between J1 and J2.

The upper dashed lines of Fig 10 represent the shield partition mentioned above. It can be made from flashing copper or thin brass.

The control box, a sloping-panel utility cabinet measuring 4 × 5 inches, houses S1, S2 and the meter, M1. Four-conductor shielded cable—the shield serving as the common lead—is used to join the two pieces. There is no reason the entire instrument cannot be housed in one container, but it is sometimes awkward to have coaxial cables attached to a unit that occupies a prominent place in the operating position. When built as shown, the two-piece instrument permits the RF pickup head to be concealed behind the transmitter, while the control head can be mounted where it is accessible to the operator.

Adjustment

Perhaps the most difficult task faced by the constructor

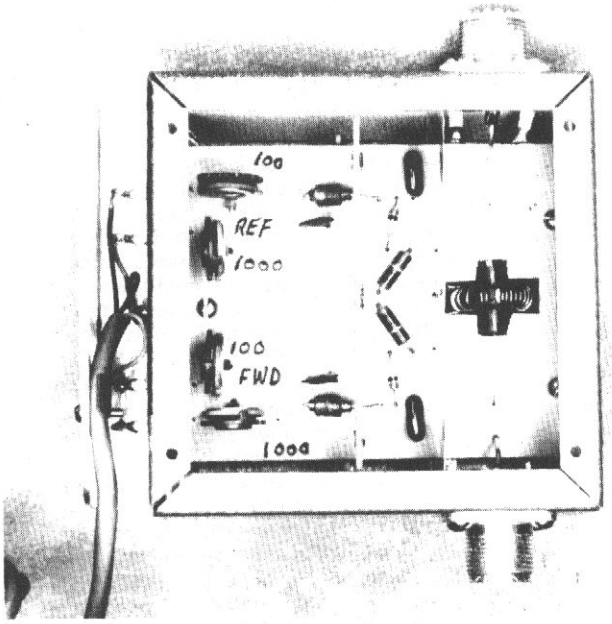


Fig 11—Top view of the RF head for the circuit of Fig 10. A flashing-copper shield isolates the primary RF line and T1 from the rest of the circuit. The second shield (thicker) is not required and can be eliminated from the circuit. If a 2000-watt scale is desired, fixed-value resistors of approximately 22 kΩ can be connected in series with high-range printed-circuit controls. Instead, the 25-kΩ controls shown here can be replaced by 50-kΩ units.

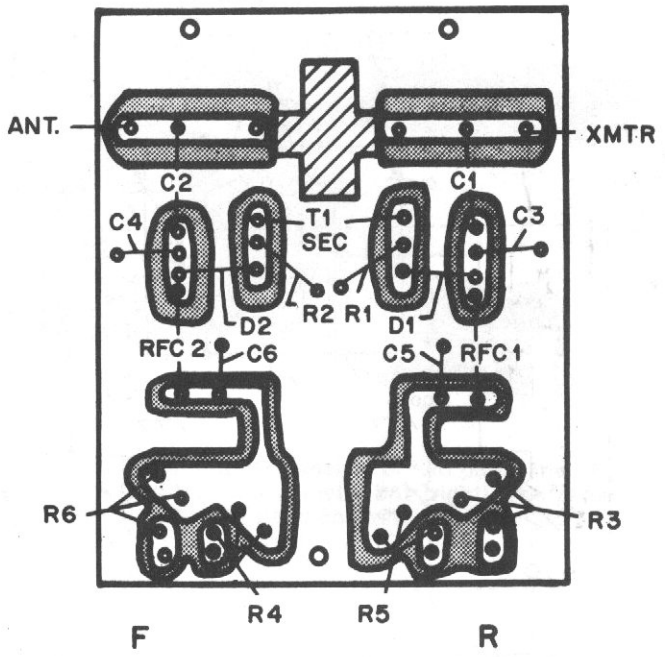


Fig 12—Etching pattern and parts layout for the RF wattmeter, as viewed from the foil side of the board. The etched-away portions of the foil are shown as darkened areas in this drawing. The area with diagonal lines is to be cut out for the mounting of T1.

is that of calibrating the power meter for a desired wattage range. The least involved method is to use a commercial wattmeter as a standard. If one is not available, the power output of the test transmitter can be computed by means of an RF ammeter in series with a 52-ohm dummy load, using the standard formula, $P = I^2R$. The calibration chart of Fig 10 is representative, but the actual calibration of a particular instrument will depend on the diodes used at D1 and D2. Frequently, individual scales are required for the two power ranges.

Connect a noninductive 52-ohm dummy load to J2. A Heath Cantenna or similar load will serve nicely for adjustment purposes. Place S2 in the FORWARD position, and set S1 for the 100-watt range. An RF ammeter or calibrated power meter should be connected between J2 and the dummy load during the tests, to provide power calibration points against which to plot the scale of M1. Apply transmitter output power to J1, gradually, until M1 begins to deflect upward. Increase the transmitter power and adjust R4 so that a full-scale meter reading occurs when 100 watts is indicated on the RF ammeter or other standard in use. Next, switch S2 to REFLECTED and turn the transmitter off. Temporarily short across R3, turn the transmitter on, and gradually increase power until a meter reading is noted. With an insulated screwdriver, adjust C2 for a null in the meter reading.

The next step is to reverse the coax-connections to J1 and J2. Place S2 in the REFLECTED position and apply transmitter power until the meter reads full scale at 100 watts output. In this mode the REFLECTED position actually reads forward power because the bridge is reversed. Calibrating resistance R3 is set to obtain 100 watts full scale during this adjustment. Now, switch S2 to FORWARD and temporarily place a short across R4. Adjust C1 for a null reading on M1. Repeat the foregoing steps until no further improvement can be obtained. It will not be necessary to repeat the nulling adjustments on the 1000-watt range, but R5 and R6 will have to be adjusted to provide a full-scale meter reading at 1000 watts. If insufficient meter deflection is available for nulling adjustments on the 100-watt range, it may be necessary to adjust C1 and C2 at some power level higher than 100 watts. If the capacitors tune through a null, but the meter will not drop all the way to zero, chances are that some RF is leaking into the bridge circuit through stray coupling. If this is the case, it may be necessary to experiment with the shielding of the through-line section of the RF head. If only a small residual reading is noted it will be of minor importance and can be ignored.

With the component values given in Fig 10, the meter readings track for both power ranges. That is, the 10-watt level on the 100-watt range and the 100-watt point on the 1000-watt range fall at the same place on the meter scale, and so on. This no doubt results from the fact that the diodes are conducting in the most linear portion of their curve. Ordinarily, this desirable condition does not exist, making it necessary to plot separate scales for the different power ranges.

Tests indicate that the SWR caused by insertion of the

power meter in the transmission line is negligible. It was checked at 28 MHz and no reflected power could be noted on a commercially built RF wattmeter. Similarly, the insertion loss was so low that it could not be measured with ordinary instruments.

Operation

It should be remembered that when the bridge is used in a mismatched feed line that has not been properly matched at the antenna, a reflected-power reading will result. The reflected power must be subtracted from the forward power to obtain the actual power output. If the instrument is calibrated for, say, a 52-ohm line, the calibration will not hold for other values of line Z_0.

If the instrument is to be used for determining SWR, the reflected-to-forward power ratio can easily be converted into the corresponding voltage ratio for use in Eq 2 given earlier. Since power is proportional to voltage squared, the normalized formula becomes

$$SWR = \frac{1 + \sqrt{k}}{1 - \sqrt{k}}$$

where k is the ratio of reflected power to forward power. The *power* curve of Fig 8 is based on the above relationship, and may be used in place of the equation to determine the SWR.

An Inexpensive VHF Directional Coupler

Precision in-line metering devices capable of reading forward and reflected power over a wide range of frequencies are very useful in amateur VHF and UHF work, but their rather high cost puts them out of the reach of many VHF enthusiasts. The device shown in Figs 14 through 16 is an inexpensive adaptation of their basic principles. It can be made for the cost of a meter, a few small parts, and bits of copper pipe and fittings that can be found in the plumbing stocks at many hardware stores.

Construction

The sampler consists of a short section of handmade coaxial line, in this instance, of 52-ohms impedance, with a reversible probe coupled to it. A small pickup loop built into the probe is terminated with a resistor at one end and a diode at the other. The resistor matches the impedance of the loop, not the impedance of the line section. Energy picked up by the loop is rectified by the diode, and the resultant current is fed to a meter equipped with a calibration control.

The principal metal parts of the device are a brass plumbing T, a pipe cap, short pieces of 3/4-inch-ID and

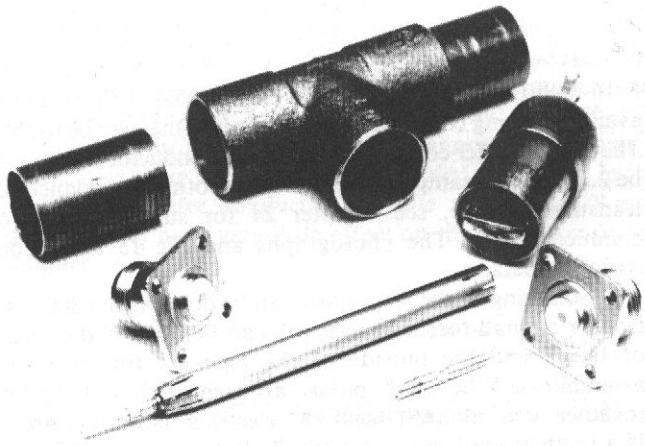


Fig 14—Major components of the line sampler. The brass T and two end sections are at the upper left in this picture. A completed probe assembly is at the right. The N connectors have their center pins removed. The pins are shown with one inserted in the left end of the inner conductor and the other lying in the right foreground.

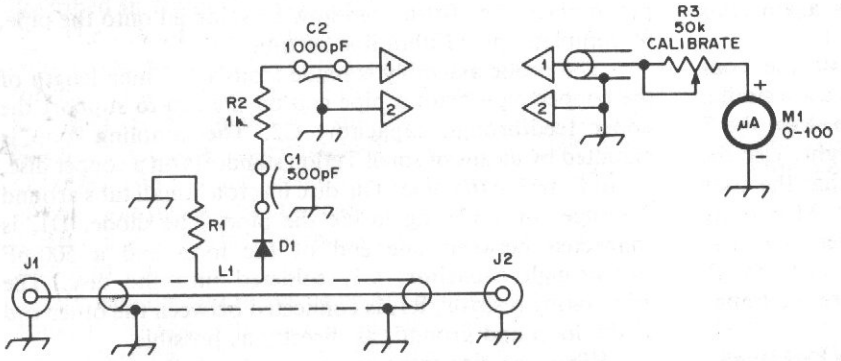


Fig 13—Circuit diagram for the line sampler.
C1—500-pF feedthrough capacitor, solder-in type.
C2—1000-pF feedthrough capacitor, threaded type.
D1—Germanium diode 1N34, 1N60, 1N270, 1N295, or similar.
J1, J2—Coaxial connector, type N (UG-58A).
L1—Pickup loop, copper strap 1 in. long × 3/16 in. wide. Bend into "C" shape with flat portion 5/8 in. long.
M1—0-100 μA meter.
R1—Composition resistor, 82 to 100 ohms. See text.
R3—50-kΩ composition control, linear taper.

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CRYSTAL SETS 2

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CRYSTAL SETS 2: SOME PRACTICAL DESIGNS

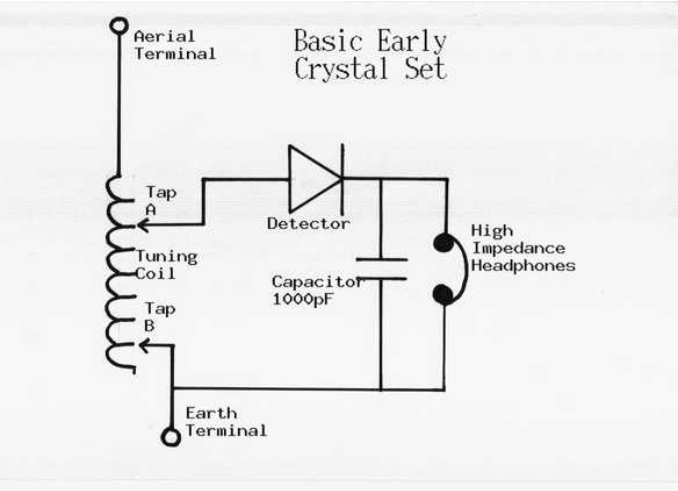
I hope that you attempt building one or two of these crystal set designs and I really do recommend that the components are carefully connected up using soldered joints onto a piece of tag-strip for reliability. However if you are new to constructing such electronic circuits then some simple solder-less techniques could be employed and these are suggested at the bottom of the page. Also see Crystal Sets Part 5 for more ideas on experimenting with crystal sets.

An early and very basic crystal set would have been nothing more than a coil of wire, perhaps 50 -100 turns, wound around a cardboard tube about 3 inches (7cm) in diameter, a detector (or cats whisker) and a pair of special High Impedance headphones (as discussed in part 1).

There would be a very large aerial strung up around the garden and the all important connection to earth.

The coil would have tapping points (connection points) at intervals of around 5 or 10 turns. See the circuit diagram on the right for details of who the set is wired together.

The tapping points on the coil allow the set to be tuned to different frequencies by adjusting the position of tap B. Tap B would be connected to the coil at differently positions by way of a crocodile clip. The fewer turns between the top (aerial end) of the coil and tap B, the shorter the wavelength received (ie the higher the frequency). Tap A would allow the detector to be connected at different positions to vary performance. There is an additional component drawn in the above diagram, the *capacitor* (value 1000pF), this is included in crystal sets that used the *High Impedance* magnetic headphones, and bypassed any remaining radio frequencies (RF) to earth.

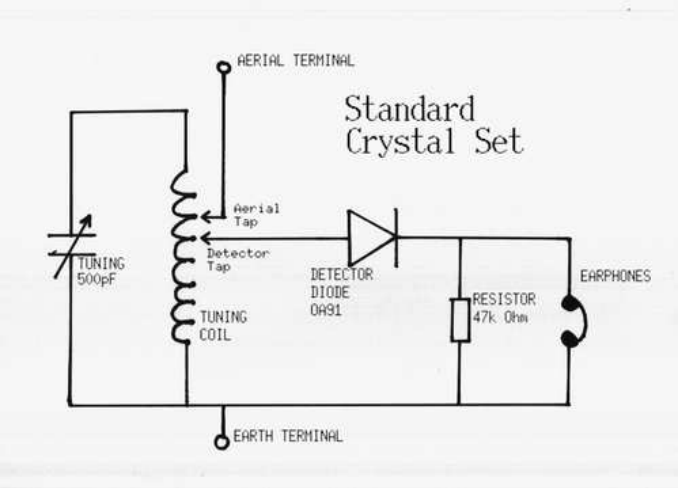


A very basic crystal set circuit.

I have not built the set described above as it is so basic. Such a crystal set above would probably have been adequate in 1920 - 1923 when there would have been only one local transmitter receivable.

When the BBC expanded transmissions and it became possible to hear more than a single station it would have became necessary to include a more convenient means of tuning the set.

This was achieved by including a Variable Tuning Capacitor, of about 500pF (0.0005uF) connected in parallel with the tuning



The Standard Crystal Set

coil forming a *tuned circuit*. The tuning capacitor would have a Bakelite knob on the spindle to aid tuning.

Because of the simplicity of crystal sets, it is often difficult to separate stations. When tuned into one station it is often possible to hear another close by station in the background, this is due to lack of *selectivity*. This can be reduced somewhat by adjusting the positions of the Aerial Tap and Detector Tap. Moving them closer to the bottom of the coil, the earthy end, reduces the load on the tuned circuit and this improves selectivity, however it does also reduce *sensitivity* which can make the station quieter. Headphones will often swamp a tuned circuit and reduce its selectivity (Q factor), so moving the tapping point lower down improves this situation. Every circumstance is bound to be different though so the best balance has to be found by experimentation. My crystal set has both the diode and the aerial connected to the same tapping point on the coil, about a quarter of the way down.

The modern 'standard crystal set' shown above uses a Crystal Earphone, since suitable high impedance magnetic headphones (of 2000 to 4000 ohms) are no longer widely available. When using a crystal earpiece the 1000pF capacitor shown in the first diagram can usually be omitted an in its place a 47k ohm resistor is connected, this ensures that the Crystal Earphone will work at its most efficient i.e. the sounds will be as loud as possible. The resistor allows DC current to flow through the circuit efficiently - this would otherwise be blocked when using a crystal earphone. In a modern crystal set the detector used is a *Diode.* Suitable diodes include OA80, OA81, OA90 OA91 and IN94 which are usually available from component stockists.

A Better Diode For Increased Efficiency

The OA47 will be of particular interest since it has the lowest forward bias voltage of any of these diodes which will make the crystal set somewhat more sensitive and therefore louder. The US equivalent of the British OA47 is the IN34.

On the right you will see my real working example of a crystal set

The large plastic knob on the front turns the variable tuning capacitor. This set receives the three UK national stations and also three local radio stations very well at my location.

There is a small 3.5mm jack socket mounted on the front of the plastic case (MB5 from Maplin Electronics) that the crystal earphone plugs into.

The coil can be seen inside the case, it is 70 turns of 30 gauge enamelled copper wire wound around the centre of a toilet roll and tapped every 10 turns, by scraping off the enamel insulation and making a small twist. The croc' clips can be seen clipped on to these twists to connect to the aerial and detector tap points.

A real working crystal set. Radio as if by magic with no battery or mains power.

THE MEDIUM WAVE COIL - MORE DETAILS

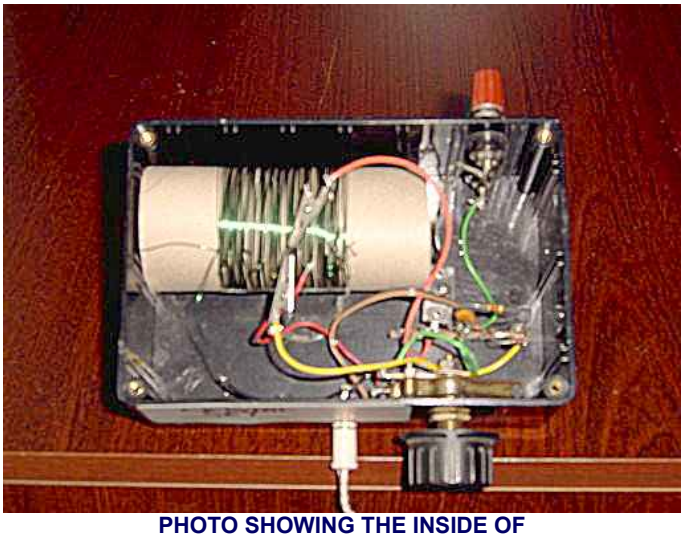


PHOTO SHOWING THE INSIDE OF THE COMPLETED CRYSTAL SET

Medium Wave Coil

The number of turns of wire required on the coil will vary depending on the size of the former (in this case the inside toilet roll) and the thickness of the wire. So to obtain the correct coverage of the medium wave band may need a little experimentation.

I usually find that between 50 to 90 turns is right and I generally use enamelled copper wire that is between 30 s.w.g. and 26 s.w.g (i.e. 0.315mm and 0.45mm diameter), so it's best to start with too many turns and then work down.

The more turns that you use the lower the frequency range will be, i.e. too many and the coverage of the top end of medium wave around 1500 - 1600 kHz will be lost, while too few and the coverage down to 500 kHz will be lost.

It is also important that the coil former is non conducting, i.e. not metallic. It could be wood or cardboard or a short piece of PVC piping and with a diameter of between 1½ and 4 inches (4 to 15 cm) are common sizes. You could try using a ferrite rod too, see below.

This particular set has a coil wound onto a toilet roll tube which consists of 70 turns of 30 s.w.g. (0.315mm dia) enamelled copper wire tapped at every 10 turns. It also has the additional small trimmer capacitor that helps match the aerial to the tuned circuit thereby improving selectivity, see below.

USING A FERRITE ROD AS THE COIL FORMER

The aerial coil could be wound onto a ferrite rod.

A piece of 10mm diameter ferrite rod of between 3 and 6 inches long (80 to 150mm) will be most suitable and will require between 50 and 90 turns of enamelled copper wire to provide coverage of the medium wave band: First make a paper tube that is held together with sticky tape that will easily slide up and down the ferrite rod. Then wind the coil over this with the windings neatly side by side. Make tapping points every 10 or 15 turns so that the aerial and diode tapping points can be adjusted.

Adjustments to the tuning range can be made by removing some wire from the coil so it is best to start off with too many turns and then work down. Fine adjustments can be made to the completed coil by sliding it up and down the ferrite rod.

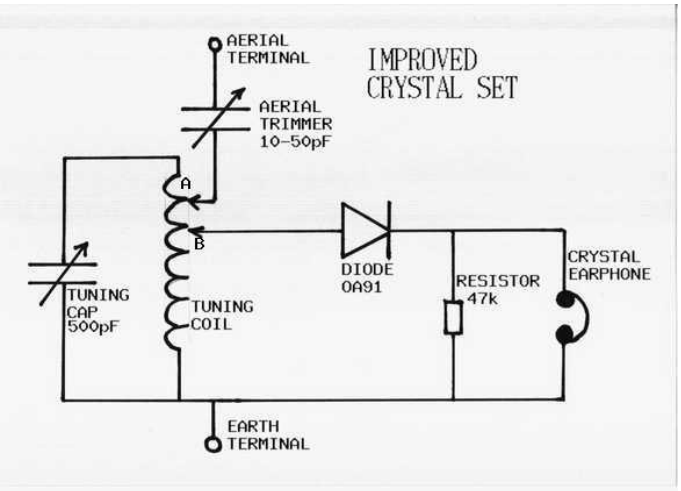
AN IMPROVEMENT TO THE DESIGN

The crystal set above also has one small, but significant, improvement over the standard crystal set and that is an Aerial Trimmer. A trimmer is a variable capacitor, very similar to the tuning capacitor, except smaller and adjusted with a screwdriver.

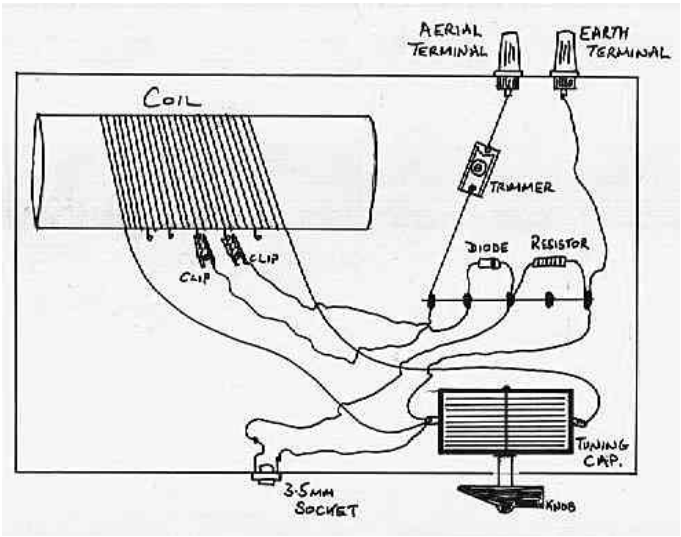
The value of the trimmer is usually around 10 - 50pF, but if a small tuning capacitor is available that will probably be just as effective. In the absence of such a variable capacitor, individual fixed ceramic capacitors of e.g. 10pF, 50pF and 100pF can be tried in this position to judge which gives the best results with the particular aerial being used.

The trimmer capacitor adjusts the coupling to the tuned circuit, reducing the load of the aerial on the tuned circuit will improve

the selectivity (Q), and it will be easier to separate stations. Again tapping points are used and I find this to be an excellent arrangement.



Improved Crystal Set design, with good selectivity



Layout Of The Crystal Set - Although this is soldered together an alternative to tagstrip would be a 5amp mains connector block so that components can be trapped in place with screws. See article below.

The picture on the right shows the general layout of the crystal set above. The coil is of approximately 70 turns is wound on the centre of a toilet roll, and has tapping points at 10 turn intervals.

The trimmer is soldered between the Aerial terminal and the piece of 5-way tag strip, and a wire goes from there to a croc' clip which is clipped onto a tap on the coil. The Diode is also soldered onto the tag strip, one end connected to a piece of wire going to a second croc' clip & connected to a tapping point on the coil, the other end of the diode is connected to the 3.5mm jack socket that the Crystal Earphone plugs into.

The 47k resistor is also connected to the earphone end of the diode and goes to earth, the earth terminal wire is soldered to the tag strip at this point too. The tuning capacitor has two terminals, one connected to each end of the coil, and one of them is also connected to earth as shown. [Where the wires cross over in the diagram, they do not touch and are not connected together].

LONG WAVES

In most areas around Europe and certainly around much of the UK you will be able to hear a Long Wave station. To receive Long Wave on a crystal set will require an aerial coil with a greater number of turns to increase its inductance.

As a good general guide a coil wound on a piece of 10mm diameter ferrite rod will require about 250 turns of enamelled copper wire: First make a paper tube that is held together with sticky tape that will slide up and down the ferrite rod. Then wind the 250 turn coil over this, the windings will have to be made over the top of each other. Make tapping points at, say, 50, 75 and 100 turns to tap the aerial and diode to.

As with the medium wave ferrite rod aerial, adjustments to the tuning range can be made by adding or removing some wire from the coil, and fine adjustments can be made to the completed coil by sliding it up and down the ferrite rod. The longer the ferrite rod the better and anything between 3 and 6 inches long (80 to 150mm) will be very good.

SHORT WAVES

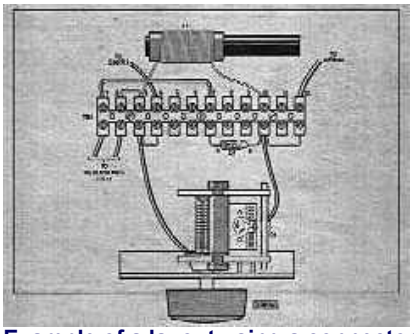
If you like experimenting, then reducing the number of turns on the coil to say 10 to 30 will allow reception of the higher frequencies, the Short Waves. I have found

that winding the coil around a 'ferrite rod' often works even better with short wave reception.

Obtain a ferrite rod about 7 to 15 cm long and about 1cm in diameter. Make a couple of small tubes of card, about 4cm long, that will fit tightly over the rod.

On one tube wind two coils using 0.5mm diameter enamelled copper wire - one coil of about 30 turns and a second one of 2 or 3 turns wound over the top of the first. Secure the windings in place with Sellotape.

On the other card tube wind a similar coil, but use about 15 turns for the first coil and for the second coil wind about 3 to 4 turns over the top, and secure with Sellotape tape.



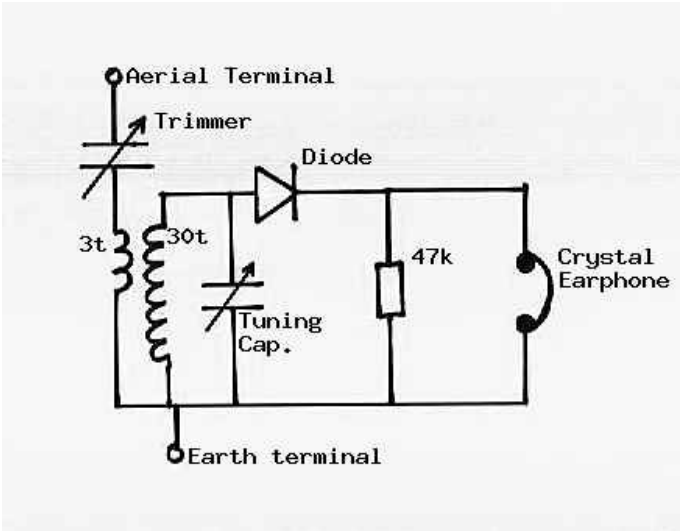
Example of a layout using a connector block to wire up a crystal set

These coils will provide coverage of short wave in two bands using the first coil for the longer wavelengths, typically 60 to 31 metre bands and the second coil for the shorter wavelengths typically 25 to 19 metre band. Wire up the circuit as shown in the circuit diagram below.

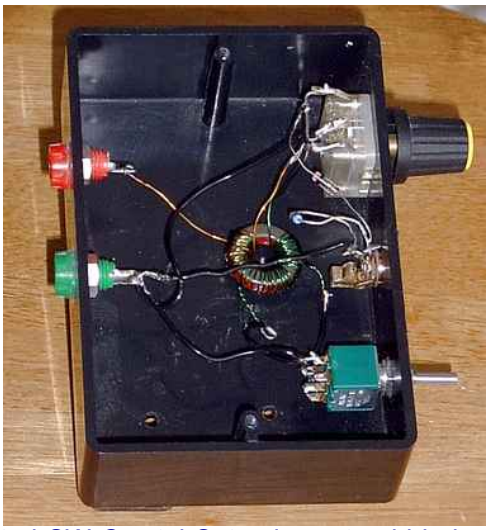
USING A TOROID INDUCTOR FOR SHORT WAVES

Even better selectivity performance can be achieved by winding the inductors (coils) on a ferrite toroids (T50-2 yellow, or green will do). The aerial trimmer need not be used if selectivity and sensitivity is found to be adequate. It's all about experimenting, and I find it best to use a trimmer or small coupling capacitor to obtain the best selectivity.

Up to 30 turns of 0.5mm enamelled copper wire can be used for the longer short waves below 10 MHz, while a winding of around 15 turns will provide coverage of the shorter short waves above 10MHz.



The circuit diagram of the Short Wave Crystal Set



A completed SW Crystal Set using a toroid inductor. Note: the main winding has a tap to allow the switch to short part of the winding and thereby give two ranges.

AUSTRALIAN DESIGN

Moving back to the Medium Waves, here is a circuit for a very interesting Australian design that promises extremely good station separation (selectivity), and having built it I can vouch for that claim, it's really excellent.

I receive three national stations and three local stations at my location with excellent clarity using a modest antenna and standard crystal earphone.

The coil is different to the other crystal sets described above, it is

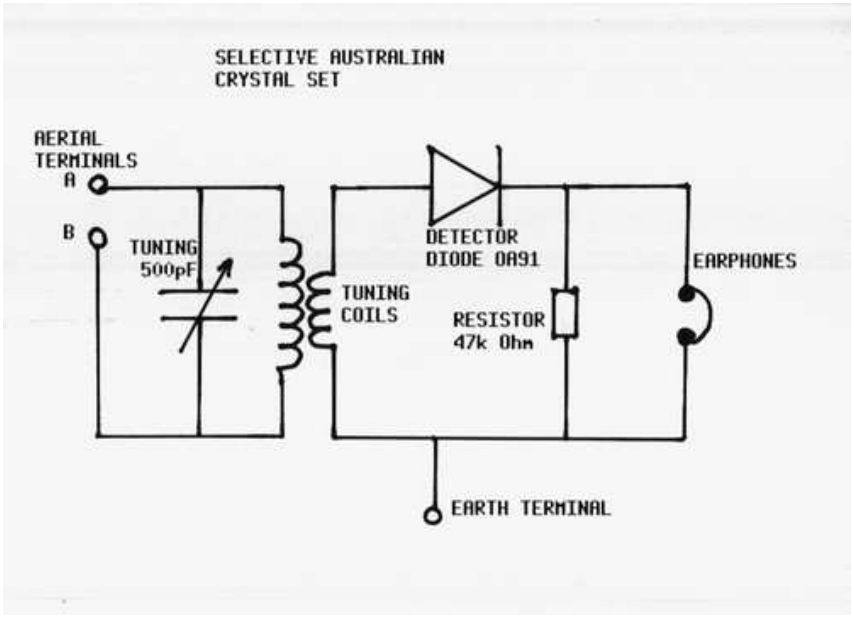
The front panel of the Australian Crystal Set

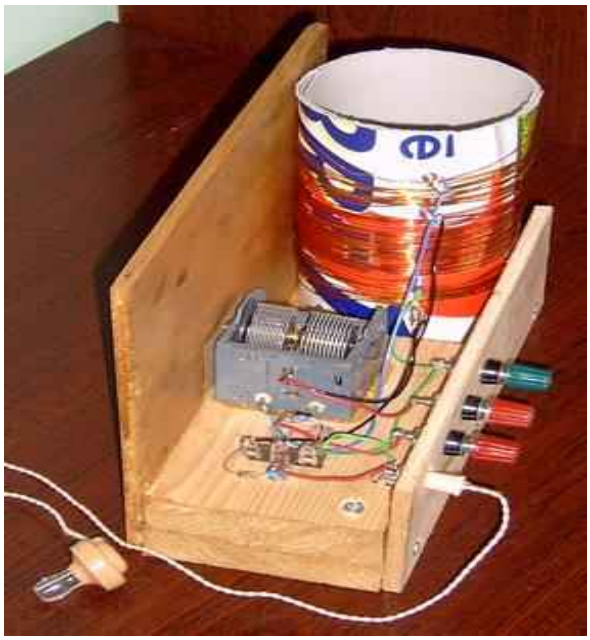
much bigger at 3¼ inches (8cm) diameter and 5inches (12cm) long. I made my coil former out of the cardboard from a breakfast cereal box - just like Blue Peter!

The design is often referred to as The Mystery Crystal Set, by Proton.

Two distinct coils are wound on it, the first one consists of about 50 turns of 24 s.w.g (approx) enamelled copper wire. The second coil is 25 turns, very close wound right over the top of the first coil using 30 s.w.g. (approx) wire, try to get this second coil wound in between the windings of the first, for better inductive coupling.

Then carefully wire up the set according to the diagram. Notice that the tuned circuit is not connected to earth and has no *direct* connection to the detector circuit. The detector circuit *is* connected to earth however. The two aerial terminals offer alternative selectivity performance, terminal A gives very good selectivity while B is very wide. I never bother with B.





Make the coil carefully and wire up this crystal set according to the circuit diagram opposite and you will be rewarded with a really high performance crystal set of a type that was used in the very early days of broadcasting in 1930's in Australia.

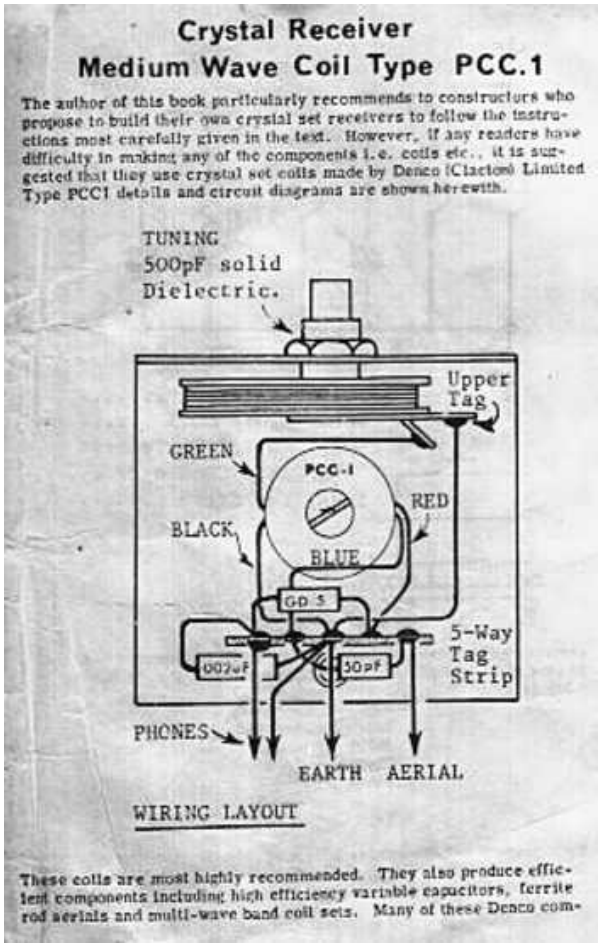
This is probably my favourite crystal set!

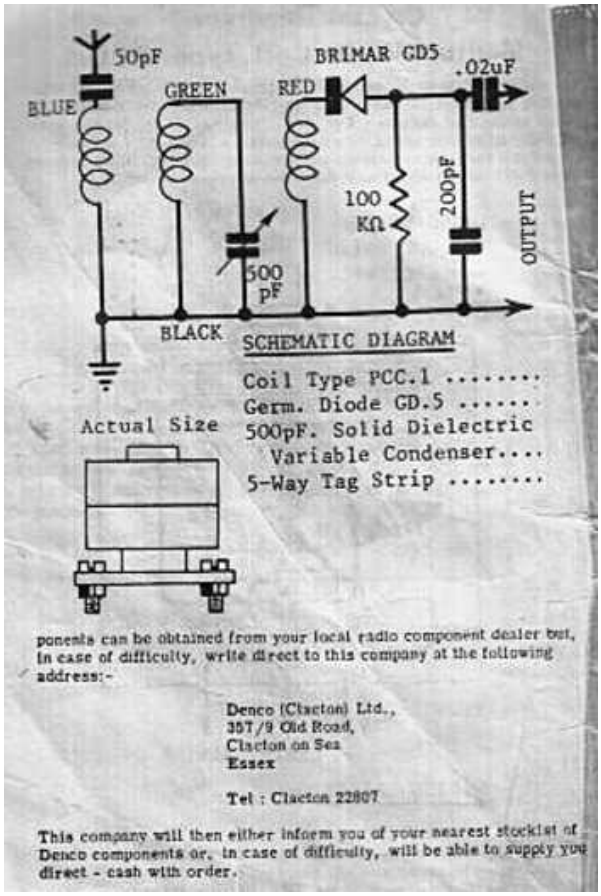
THE DENCO PCC1 COIL

The PPC1 coil was a commercially manufactured by Denco Clacton Ltd and was popular among hobbyists not keen on going to the bother of winding their own fiddly little coils. As a child I wanted try one of these coils and sent away for one by mail order. It arrived a few days later in a little cloth bag, like a miniature pump bag, with protective wrapping inside.

The coil windings are entirely enclosed in what I can only describe as a cylindrical ferrite 'shell', the four very thin connecting wires exiting, two either side, from small apertures in the 'shell'. The performance of the circuit shown below I seem to remember was quite pleasing. Unfortunately I cannot find the set or the PPC1 coil at the moment, but here is

a reproduction of the circuit diagram and data:



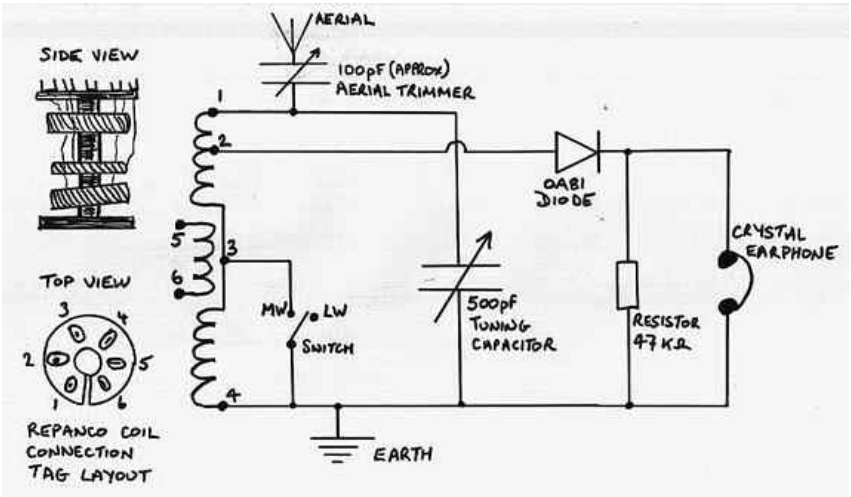


THE REPANCO DRR2 COIL

I recently rediscovered an old Repanco DRR2 Longwave / Mediumwave coil that must have been kicking around in my junk box since the 1970's.

The DRR2 coil was made by Repanco in Coventry. It came with a page of suggested circuit diagrams which I thought had been lost to the mists of time, but it recently came to light again, so I have now copied it below.

Once again I included an aerial trimmer which can be adjusted to improve selectivity.



The circuit diagram of the crystal set using the Repanco DRR2 coil

Repanco Ltd was formed by two ex-army signals engineers and from the earliest days of radio supplied crystal set kits and coils to radio construction enthusiasts.

The Repanco DRR2 coil was for medium wave and long wave intended for use in when building simple crystal set and valve radio

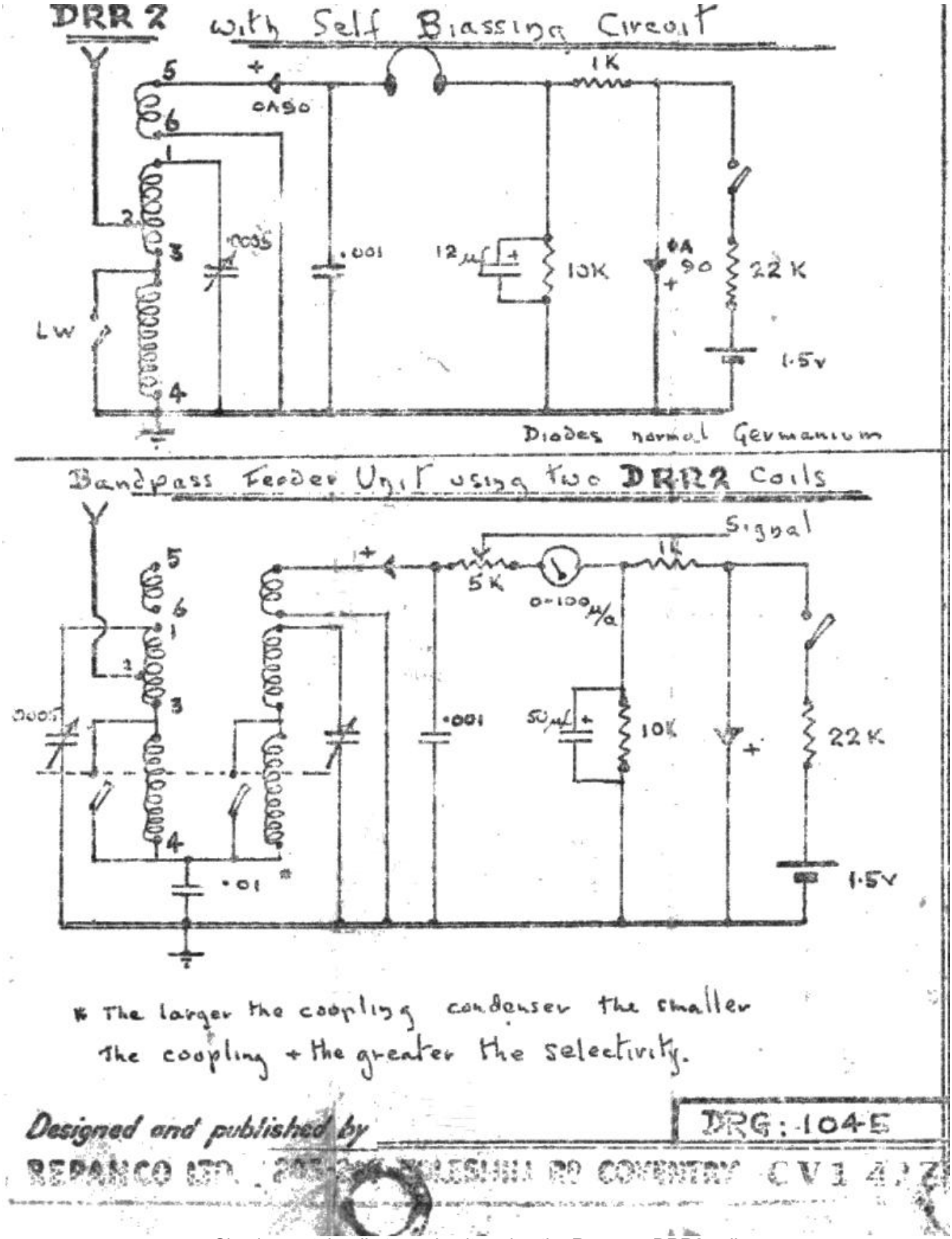
A 'lash-up' of the Repanco crystal set

circuits.

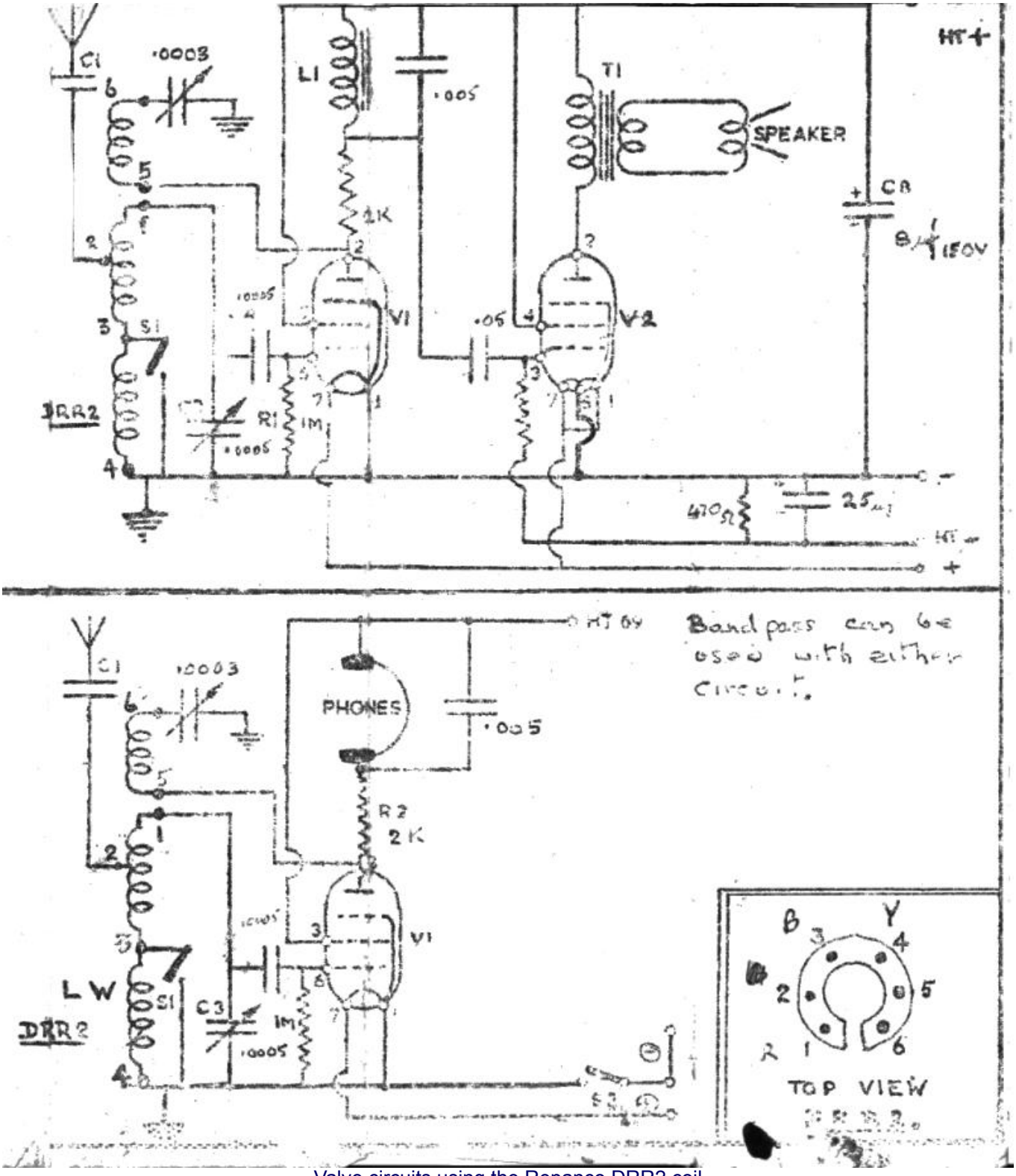
It consists of three coils; a Medium Wave coil at the top that includes a tapping point (for the aerial); a coupling coil or tickler in the middle; a lower coil which can be connected in series with to top coil to provide Long Wave reception.

I have built a quick crystal set with the coil and it provides good reception with excellent selectivity, so it must have a very good Q factor.

The Repanco DDR2 coil was provided with a simple Foolscap size information sheet that showed four different radio circuits. Sadly the sheet does not give a huge amount of information and my copy is now rather tatty and faded - it is copied below:



Simple crystal radio type circuits using the Repanco DRR2 coil



Valve circuits using the Repanco DRR2 coil

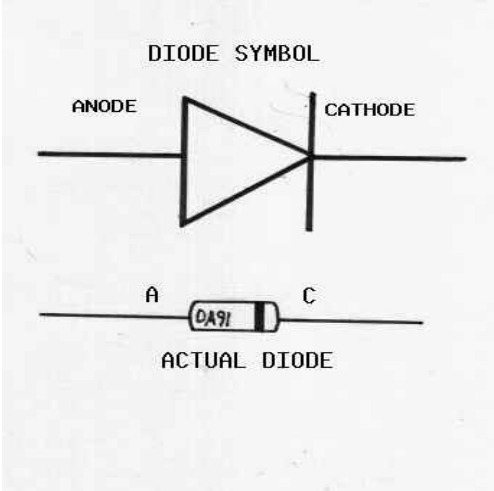
Repanco Ltd no longer produces radio coils and crystal set kits for the radio construction enthusiast, as it did in the early days of radio. In 1986 it was renamed Repanco Bartlett Ltd when it merged with Bartlett Electronics. The company moved from the Foleshill Road to Unit 24, Albion Industrial Estate, Endemere Road, Coventry CV6 5NT and now specialises in transformers and wound components and can design and manufacture to commercial customer requirements, their website is: http://www.repancobartlett.co.uk/

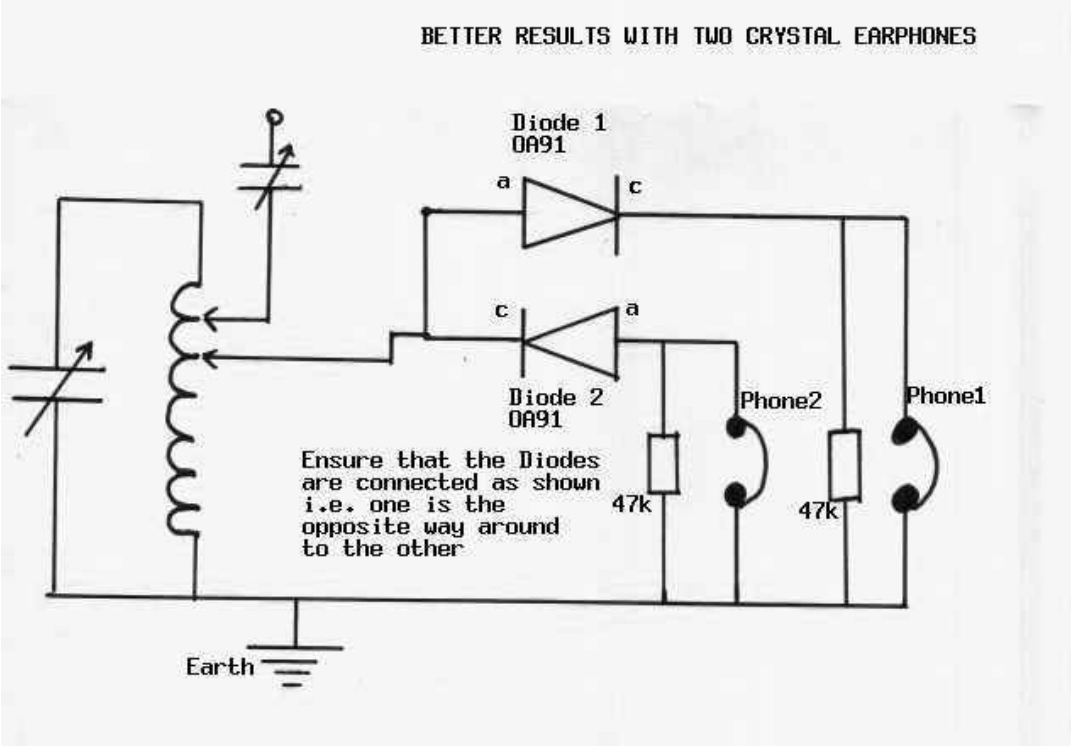
CRYSTAL EARPHONES

Here is a good idea and well worth trying, to maximise the use of sound output from your crystal set why not use dual crystal earphones? Having an earphone in each ear helps to block out extraneous noises helping the listener to better concentrate on any weaker stations received.

Using the circuit below, one earphone makes use of one half cycle of the radio wave while the second earphone uses the other half cycle of the wave that would have previously gone to waste when using just one diode. Ensure that the diodes are connected up according to the diagram i.e. one diode is connected the opposite way round to the other.

Also try to make sure that the diodes and crystal earphones are similar to obtain the best results. (You could simply connect two crystal earphones to the same terminals of the single diode, but this would not be as efficient and the sounds would be much quieter.)





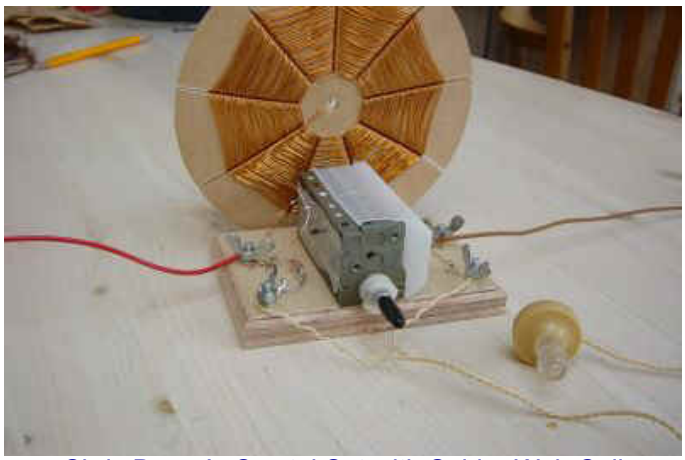
A note about Crystal Earphones: It will be worthwhile buying several different ones from different sources as performance varies between manufacturers quite markedly. I have found the ones marked 'Japan' on the back are the most sensitive and therefore loudest, whereas ones marked 'Receiver' 'Taiwan' are often a little less sensitive and therefore quieter and sometimes more 'tinny' sounding.

As mentioned previously it has been noted that the OA47 diode will be of particular interest since it has the lowest forward bias voltage of any of the common diodes available. This will make the crystal set somewhat more sensitive and therefore louder. The US equivalent of the British OA47 is the IN34.

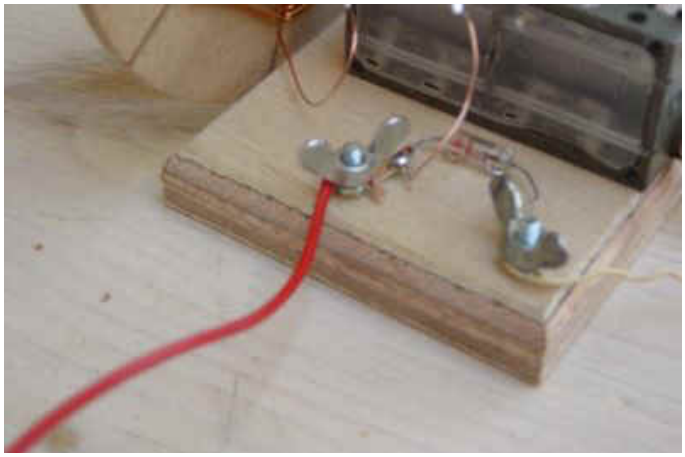
SPIDERS WEB' COIL

Here is an interesting concept sent in by Chris Dorna of the Vught North Scouts in the Netherlands. It is a crystal set made out of a coil wound in the form of a spiders web:

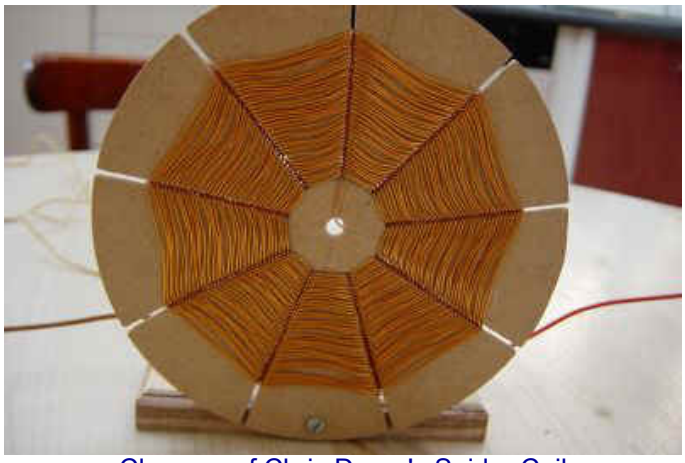
See more HERE.



Chris Dorna's Crystal Set with Spider Web Coil



Detail of the germanium diode



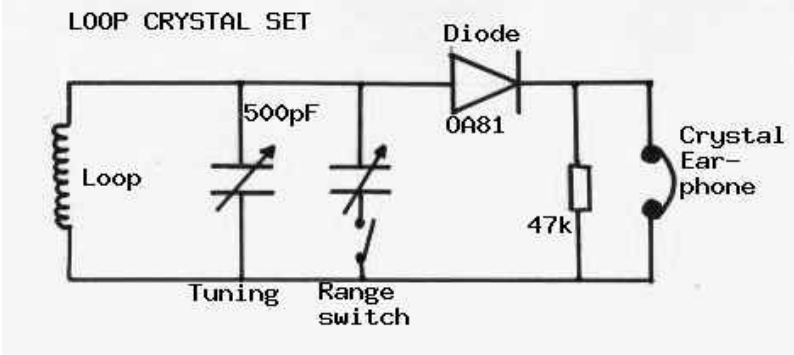
Close up of Chris Dorna's Spider Coil

LOOP THE LOOP!

A crystal set can also be made that does not need a large long wire aerial. If you have ever made a loop aerial for medium wave or long wave DX-ing, then it is a simple matter to add a diode, resistor and a socket to connect a crystal earphone that will allow reception of nearby stations.

See my section on Loop Aerials and ATU's for more constructional details.

A portable loop aerial, that incorporates a crystal set



The circuit diagram of the Loop Crystal Set. The loop is 10 turns of 7/0.2mm 'hook-up' wire wound on a 40cm (17") former made of attractive plastic edging strip available from many DIY stores. The loop is very directional in its pick up pattern, which can help eliminate interference from some stations by rotating the loop. The switch and additional capacitor allow tuning of the lower medium wave band from about 650 to 520 kHz. Having a loop with 50 to 60 turns of wire will tune into the Long Wave band.

DIODES - For Crystal Set Use - some notes by Felix Scerri

Germanium diodes for crystal set use.

Although I'm a fan of these new silicon schottky BAT 46 diodes, good germanium diodes still have a lot to offer, especially in terms of 'weak signal' sensitivity. Last night I did an experiment.

I sorted through quite a few of my hundreds of acquired random germanium diodes looking for particularly 'sensitive' ones. I tested this by tuning in a weak AM station and comparing the detected DC output level and also the apparent 'loudness' of the audio signal.

Even amongst germanium diodes of the same type, there was enormous variation all the way from excellent to poor! For very weak signals, germanium diodes 'detect' in the 'square law' region below the diode conduction 'knee', in a rather different part of the curve than with much stronger signals (way beyond the diode knee).

When testing germanium diodes for weak signal sensitivity, the inherent capacitance of the diodes is also a factor, and the 'tuning' may change somewhat and will need to be readjusted with every diode tested!

In the end, out of a large number of germanium diodes tested, I found three or four germanium diodes with excellent weak signal sensitivity and the rest were poor. One other interesting thing, good germanium diodes 'sound' different, rather more 'rounded and smoother' than the schottky's which tend to sound mercilessly clean, almost clinical. I also found almost no variation in weak signal sensitivity with my BAT 46 schottky diodes. Take your pick!

Regards, Felix Scerri VK4FUQ 14/03/2012

SOLDERLESS CONSTRUCTION IDEAS

For a novice the use of a soldering iron may seem a bit daunting at first and while the most reliable results will be obtained with a good soldered joint using a tag strip as shown below, the circuits can still be made without the use of a soldering iron.

5 WAY TAG STRIP

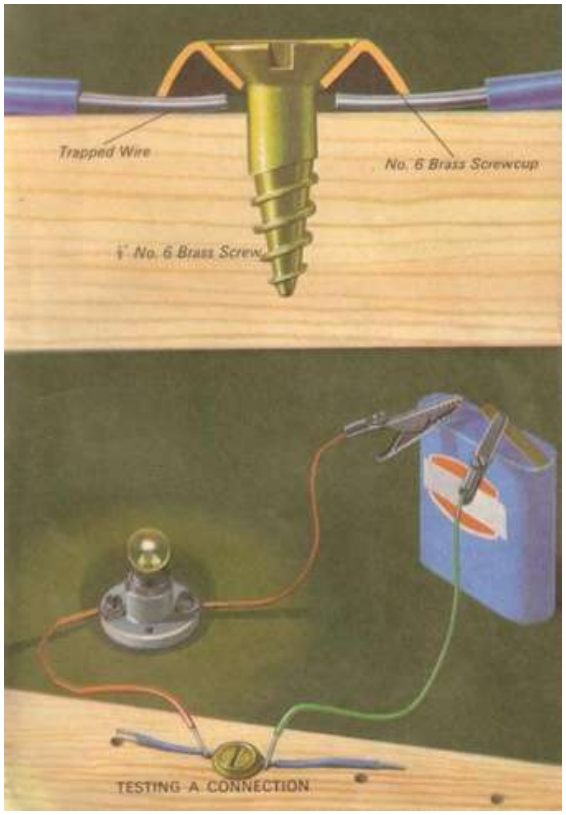
36 WAY TAG STRIP - TWO ROWS

The very simplest circuits could be wired together ,with a little ingenuity, with the component wires being held together in the grip of solderless crocodile clips, whereby the connecting hook-up wire is fixed to the croc' clip by a screw rather than solder.

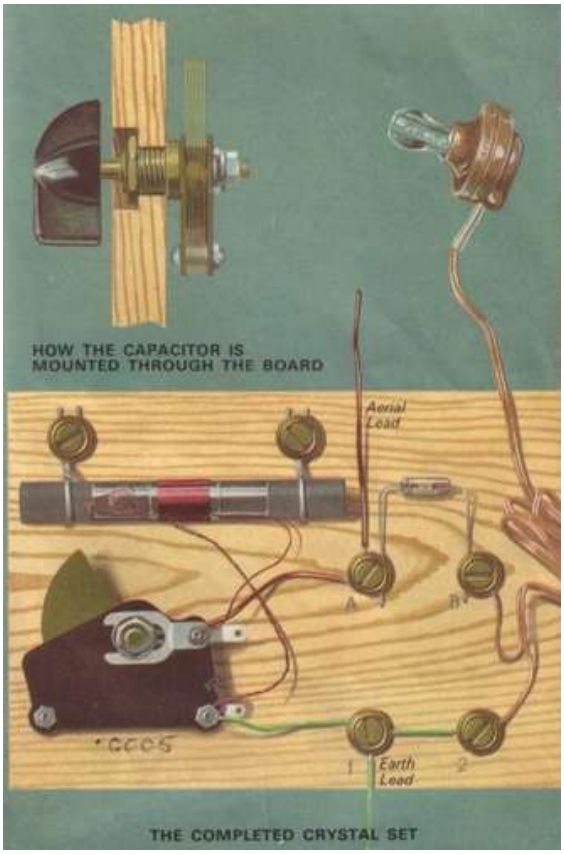
For more the slightly more complex circuits a plastic Terminal Block (sometimes referred to as a choc' or chocolate block) can be utilised very effectively indeed. These are used in mains wiring and are available in various sizes; 2 Amp, 5 Amp, 15 Amp and 30 Amp. The 5 and 15 Amp Terminal Blocks I have found to be the most suitable. The various component wires can be trapped securely with the screw at each junction point. This method also makes it easy to change the components around when experiment with different circuits. See The EXPERMENTAL CRYSTAL SET for more details in Part 5.

'CHOCOLATE' TERMINAL BLOCK

The Ladybird book called 'Making A Transistor Radio' (also shown on the TRF Radio pages) detailed a very novel approach using brass screws with screw-cups to trap the component wires at each junction point:



THE BRASS SCREW AND SCREW-CUP METHOD OF CONSTRUCTION



A VERY SIMPLE CRYSTAL SET USING THE BRASS SCREW AND CUP METHOD

Crystal Sets Part Part 3 >

No AM radio stations or transmitters in your locality or country?

Has your local medium wave broadcast station closed or been moved to VHF/FM or Digital? Don't worry. You can still build and experiment with crystal sets and TRF radios by also buying or even building a simple low power AM transmitter. So, not only can you use your crystal sets but you can also run your own radio station that can be heard in and around your home - playing the music or programmes that you want to hear!

SSTRAN AMT3000: Superb high fidelity medium wave AM transmitter kits from SSTRAN. Versions available for 10kHz spacing in the Americas (AMT3000 or AMT3000-SM) and 9kHz spacing in Europe and other areas (AMT3000-9 and AMT3000-9SM). Superb audio quality and a great and well designed little kit to build:
http://www.sstran.com/pages/products.html

http://www.sstran.com/

Other AM transmitters available:

Spitfire & Metzo: Complete, high quality ready built medium wave AM Transmitters from Vintage Components: http://www.vcomp.co.uk/index.htm Vintage Components offer a choice of the high quality Spitfire and Metzo transmitters:

SPITFIRE AM Medium Wave Transmitter with 100 milliwatt RF output power:

http://www.vcomp.co.uk/spitfire/spitfire.htm

METZO AM Medium Wave Transmitter with built in compressor:

http://www.vcomp.co.uk/metzo/metzo.htm

AM88 LP: A basic AM transmitter kit from North County Radio.
http://www.northcountryradio.com/Kitpages/am88.htm

LINKS:

BOWOOD ELECTRONICS - A friendly, helpful and very speedy source for many of your electronic components at prices that won't frighten your wallet!

THE FOXHOLE and P.O.W RADIOS - Simple crystal set receivers used by soldiers during the war and by prisoners of war (P.O.W.'s).

VINTAGE COMPONENTS - A great resource for crystal sets, components, valve radio kits and medium wave AM transmitters!

6V6 - Electronic Nostalgia and Vintage Components

Crystal Sets Part 3 >

Crystal Sets (Part1) | **Spider's Web Crystal Set (Part 3**

Crystal Set by Kenneth Rankin (Part 4) | **Experimental Crystal Sets (Part 5)** | **Crystal Radio Links**

^Top Of Page

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Building an AM radio

Objectives

By the end of this lab, the students should be able to:

- describe the function of the variable resistor, variable capacitor in the radio circuit
- describe how variable capacitors are connected in circuits
- describe how variable resistors are connected in circuits
- describe the "hidden workings" of a electronics prototyping board
- construct a AM radio (or any circuit) using a project board, components, jumper wires and a diagram

Introduction

Amplitude Modulated (AM) radio signals carry an audio signal (electrical information that can be converted to sound). The audio information is encoded on the radio wave by changing the amplitude of a high frequency radio wave. The base frequency of the electromagnetic radio wave (called the carrier wave) is much higher than the frequency of the audio information it carries (550kHz to 1600kHz versus 20Hz to 20kHz). The diagram at right shows how the encoding works. The amplitude of the carrier wave is not constant. The fast variation is the carrier wave. The relatively slow variations in the amplitude create an envelope that shows the frequencies in the audio signal. The diagram is somewhat distorted to show the principle. The amplitude of a typical radio signal varies much more slowly than the picture suggests. (How many periods of a 600kHz carrier would have to be shown in order to show half of a period of a 100 Hz sound?)

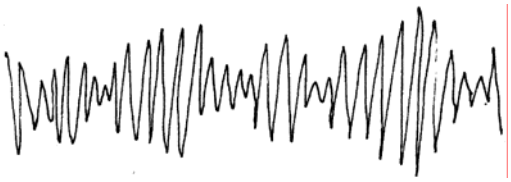


Figure 1: Amplitude modulated carrier signal

An extremely basic AM radio, consisting of a detector (a diode in the circuit at the right) and an antenna, is shown at right. The diode only allows current to flow in the direction indicated by its symbol. This circuit detects AM signals and converts them to audio signals.

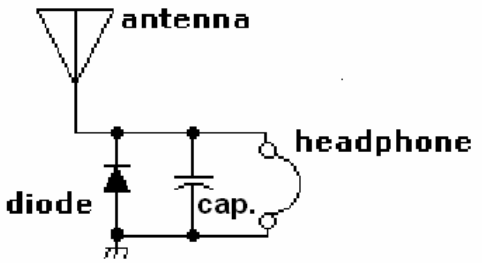


Figure 2: Simple AM detector circuit

The second circuit at the right is a (slightly) more sophisticated version. The design includes a tuning section. By setting the resonant frequency of the circuit to the frequency of the desired carrier wave, the user selects one carrier frequency for conversion.

The radio you will build (see the diagrams on the last page of the handout) includes an AM detector (a transistor in this case), a low pass filter (to remove the carrier wave from the audio signal) and an amplifier.

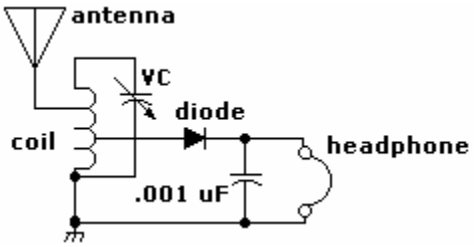


Figure 3: AM detector circuit with tuner block

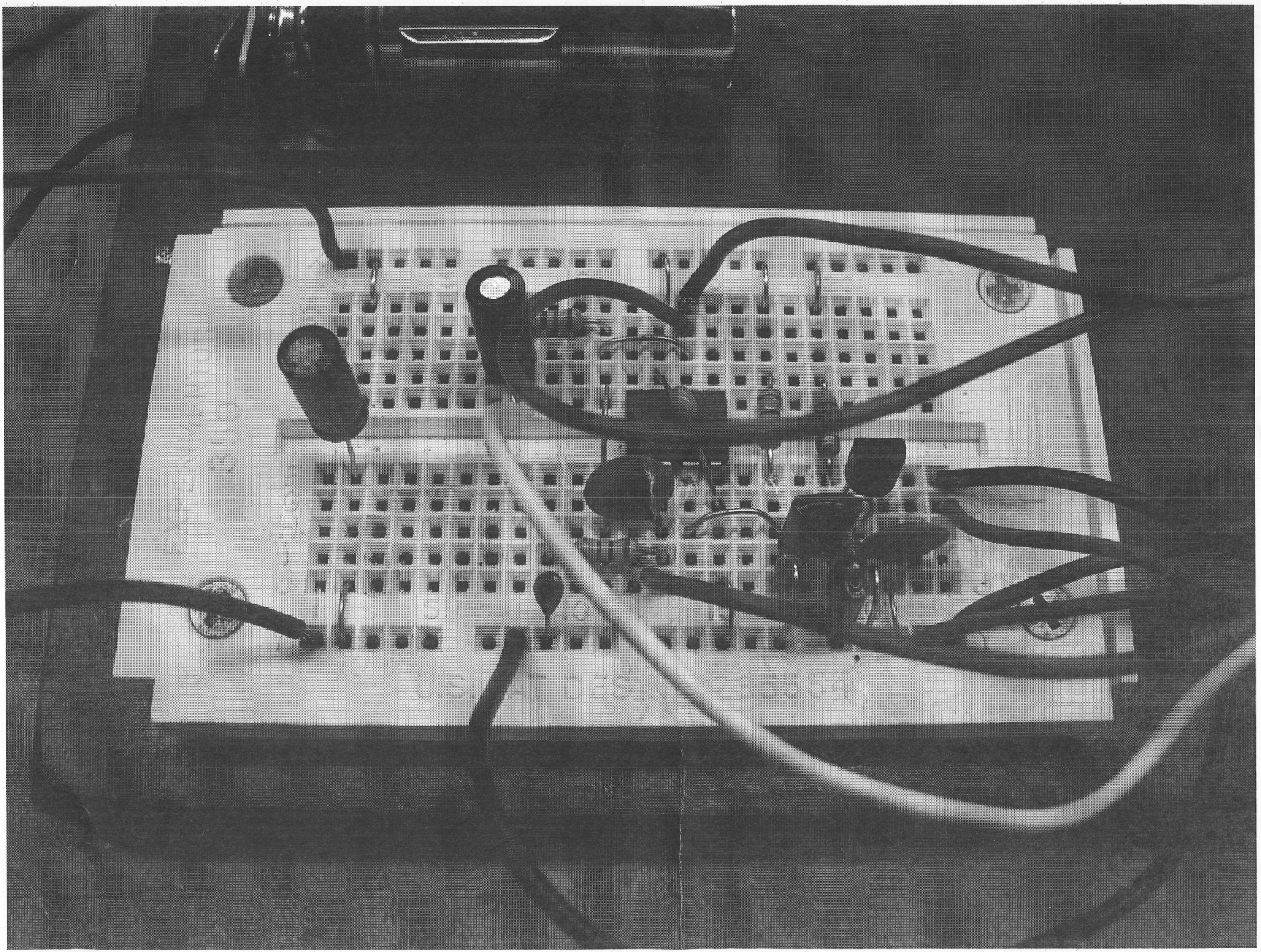
Web Resources

The web has some good resources on AM radios. Googling "crystal radio" yields quite a bit of info on how to build a radio. One such site is http://sci-toys.com/scitoys/scitoys/radio/radio.html. The web also has some good explanations of how the AM detector works. Google "AM detector" or "envelope detector". One such resource is http://www.st-andrews.ac.uk/~www_pa/Scots_Guide/RadCom/part9/page2.html.

Building an AM radio

Instructions

1. Some of the AM radio is already made. A battery holder, coil, variable capacitor, potentiometer (a.k.a. variable resistor), project board and earphone jack have been mounted to a small piece of plywood.
    - Identify which component is which.
    - Use a multimeter to figure out how the potentiometer works. (Hint: Use the ohm-meter setting on the multimeter).
2. Prototype board is used for making and testing temporary circuits. Prototype boards have internal connections. The internal connections allow you to build a circuit with relatively few wires needed and to avoid soldering). Obtain an un-mounted project board. Inspect it. Figure out the pattern of internal connections. You can do this by looking at the back of an un-mounted project board or by using an ohm-meter.
    - Which holes will be at the same potential (provided small currents are flowing)?
    - Which holes can be held at different potentials? In other words, which sets of holes are electrically isolated from each other?
3. Obtain a kit of components for your radio. Inspect them.
    - Identify the different kinds of components: resistors, capacitors, transistors, and integrated chip.
    - ***Hooking up asymmetric components incorrectly will lead to unintended operation*** (more often than not unintended operation means ***self-destruction***). Identify all non-symmetric components. Note: Some "two-legged" components, including many capacitors, are asymmetric.
4. Use the diagrams on the next page to construct the circuit. You may note slight discrepancies between the parts in your kit and the parts provided. In particular, the resistance values called for might not match the contents of your kit. In general, you can use resistance values that are within a factor of two or so and still get the circuit to work. ***Pay attention to the proper connection of asymmetric components*** (including capacitors)! ***Leave the battery out*** until you have visually double-checked the circuit against the diagram.
5. Put in the battery and test the radio. In order to get a signal, you may need to …
    - tune the radio. (Which circuit element does this?)
    - turn up the volume on the earphone (Which circuit element does this?)
    - rotate the entire radio. (Why might the orientation matter? Which element needs to be aligned? What does it need to be aligned with? Why?



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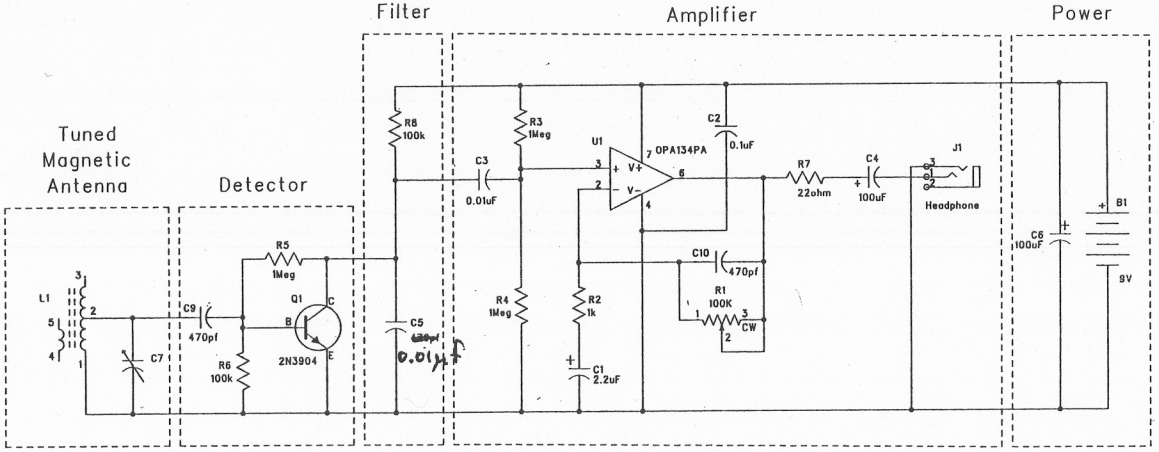


DRAWN: David McGaw	DATED: May 24, 2010	COMPANY: Dartmouth College	
CHECKED:	DATED:	TITLE: P14 AM Radio	
QUALITY CONTROL:	DATED:	CODE:	SIZE:
RELEASED:	DATED:	DRAWING NO: AM Radio	REV: 0
		SCALE:	SHEET: 1 OF 1

CONSTRUCTION OF RADIO RECEIVING APPARATUS

By The Laboratory Staff

Editor's Note: The apparatus described in this article has been constructed and tested in the Everyday Mechanics Experimental Station. The models may be seen at the Publication Offices of the magazine.

IN presenting the two designs embodied in this article, we do not wish to pose as originators of new types of apparatus, and neither do we wish to give the impression that the apparatus we have constructed is superior to anything on the market. The loading inductance presents no startling improvements in design or construction and the receiving transformer is just a good, substantial piece of apparatus that an advanced amateur need not be ashamed to construct and operate.

What we have endeavored to do, however, is this: We have tried to eliminate the defects, principally mechanical, that seem inherent in amateur apparatus; we have attempted to produce a design pleasing to the eye, convenient and smooth in operation, simple and inexpensive in construction, and of correct proportions and specifications from the radio standpoint.

The old "loose coupler" type of receiving transformer has long been a favorite for the very simple reason that it gives such universal satisfaction. The writer well remembers the experience of one of the best known manufacturers of amateur apparatus of the better grade. For years this concern had manufactured and sold great quantities of a certain number—a comparatively simple receiving transformer with sliding contacts and telescoping primary and secondary. True, the workmanship was excellent—as it was on every number turned out by the company—but there was nothing startling about that "loose coupler" to give it such a reign of popularity. However, when the manufacturer tried to substitute a perfect wonder of a tuner, with variometer coupling and a myriad of instrument-switch contacts, offering the new and improved device at the same figure as that asked for the plebeian coupler, the trade appeared to receive the newcomer with indifference and doubt. After several years of pushing and advertising, the little tuner had to be discontinued, relinquishing the field to a revival of the old-fashioned coupler. It is an actual fact that the demand for the old favorite was greater, probably just because of its simple goodness, than that for the improved instrument which was really a much better buy, value considered, than its successful predecessor.

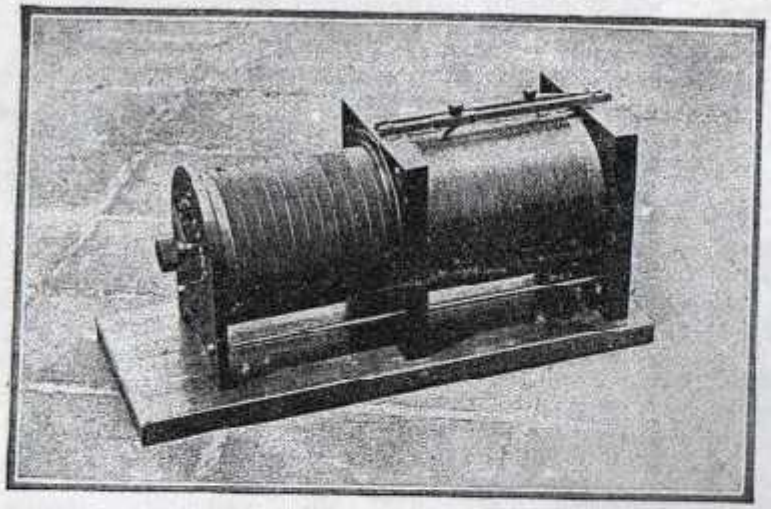


Fig. 1. Receiving transformer completed

Perhaps the reader will wonder at this digression. It is given merely to justify the publication of the design of a type that has been described many, many times in contemporary magazines. If our readers could but see our models, turn their contact knobs, and listen in to the gratifying results obtained, this justification would be unnecessary. Whatever adverse criticism this article may bring forth, the fact remains that there are probably more of these simple receiving transformers in use, and in *successful* use, too, than there are of any or all other types.

The Receiving Transformer

This transformer has instrument switch adjustment for secondary inductance and sliding contact adjustment for the primary. The construction is such that no metal (supporting rods) is used inside either primary or secondary windings. While the deleterious effect of metal in such cases is perhaps open to question, we have avoided its use as unnecessary in our design.

The transformer responds to wave-lengths as high as 2,500 meters without the loading inductance. This is believed to be the "happy medium" which reduces objectionable "dead ends" to a minimum and still affords a wave-

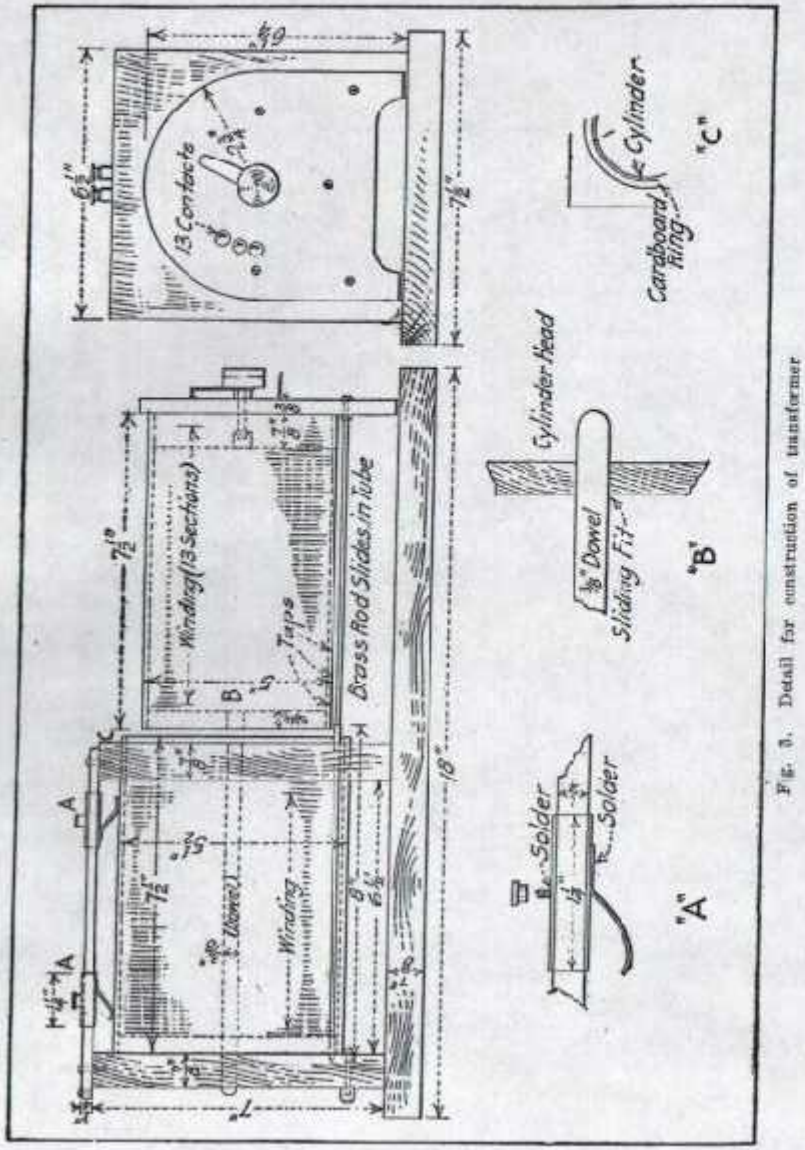


Fig. 3. Detail for construction of transformer

length of great utility. The figure quoted is based upon the use of the instrument with the average antenna of the advanced amateur.

The construction involves some careful but not necessarily difficult wood-working. A lathe is not at all necessary, although it is, as always, desirable. If a jig saw is available, most of the turning can be avoided. The only real, good excuse for a lathe is in turning the wooden heads that fit into the cardboard cylinders, and also in winding the cylinders. The latter operation may readily be done, however, by mounting the cylinders between centers on a base board. Another effective substitute for the lathe in both turning and winding operations is an ordinary polishing head that can be purchased for a couple of dollars. The discs may be mounted on the taper thread and a very presentable job of turning done with the broken end of a flat file held on a simple rest.

The jig saw, however, is an essential. If one of the foot-power variety is not available, the hand fret saw frame will answer, although in our experience there is no comparison between the two.

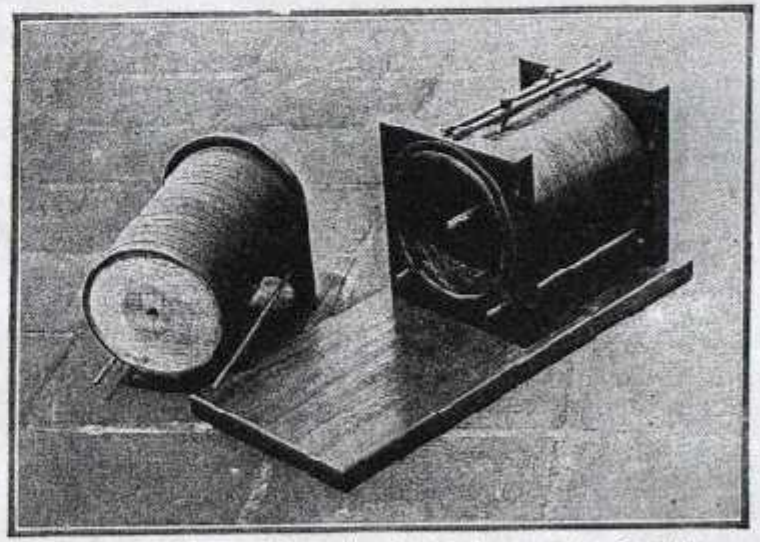


Fig. 2. Receiving transformer showing dowel arrangement for sliding

The woodwork had best be done first of all. Figs. 1, 2 and 3 will give a good idea of the appearance and construction of the in-

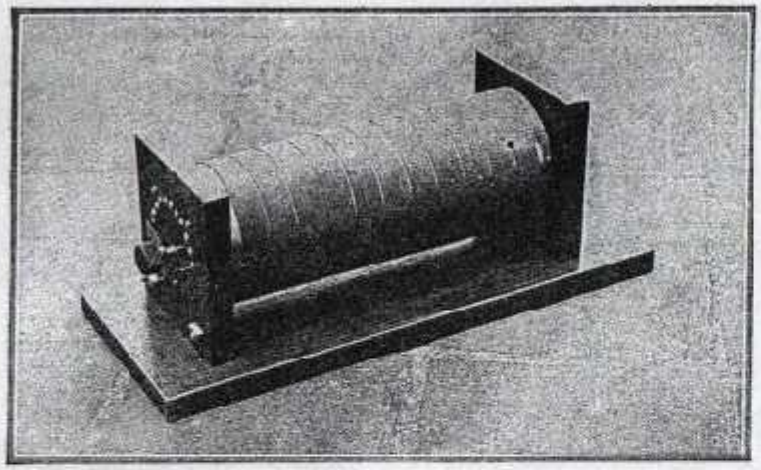


Fig. 4. Loading coil completed

cates, the secondary support is gripped between nuts and washers on the threaded ends of the rods.

We are now ready for the winding. The primary is wound directly upon a cardboard cylinder 5¼ in. diameter and 7⅛ in. long. The winding is No. 24 bare copper wire. The winding is done with two wires wound in parallel and then, after the ends are carefully secured by pushing through holes in the cylinder, and plugging with wooden pegs, one wire may be unwound. This will leave a firm, neat winding of bare copper wire with each turn separated from its neighbor by the thickness of the wire. The winding should be made about 5½ in. long, leaving ½ in. of space at either end of the cylinder when the latter is mounted in the wooden supports. After the winding is finished, it should be given two generous coats of shellac, taking care to see that the fluid runs down between turns in order that they may be sealed to the surface of the cylinder. The contact path for each slider is to be scraped in the winding after the shellac is quite hard and dry. This operation, however, need not be attempted until the cylinder is mounted and the sliders are in place on their rods.

The secondary winding consists of a single layer of No. 28 S. S. C. magnet wire wound in 13 sections upon a cardboard cylinder 5 in. diameter and 7½ in. long. The winding amounts to a single layer

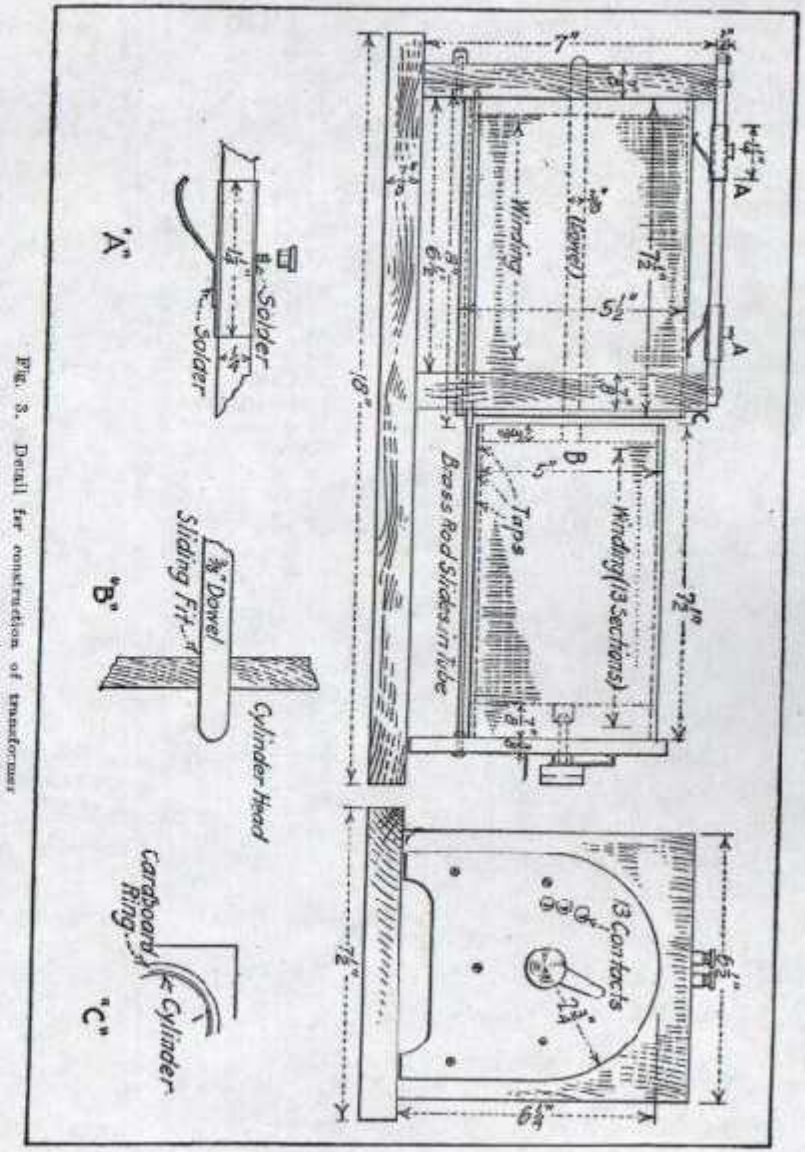


Fig. 3. Detail for construction of transformer

which is wound in the following manner: The winding is started from one end and continued for ½ in., at which point the hand guiding the wire is jumped to a distance which leaves a space of about 1-16 in. between the finished turn of the section just wound and the beginning of the next section. At the point where the "jump" is made a pin-hole is punched into the cardboard cylinder. Through this hole the tap which leads to the contact point will be made. This process is to be continued until 13 sections have been wound, at which point about 6½ in. of the surface of the cylinder will have been covered with the wire. Care must be taken to see that the holes through which the taps are taken are in a line, which will, of course, be at the bottom of the cylinder when it is mounted on the secondary support.

There are a number of ways of securing the silk-covered wire to the cardboard cylinder, but the best method in our experience is first to coat the cylinder with two or three applications of shellac, allowing the varnish to become "tacky" before placing the winding. This method will firmly secure each turn to the cylinder, and at the same time will obviate the necessity for shellacing the silk after the winding is complete.

The taps are made with lengths of slightly heavier bare copper wire pushed through the holes in the cylinder and soldered with a very fine copper, one to each of the "cross over" wires between sections. These taps should be long enough to extend through the ⅞ in. head to make contact with the points upon which the instrument switch bears. Each tap is to be covered with a sleeve of fine rubber tubing to prevent possible short circuits inside the cylinder.

The next operation, before going further with the taps, is to insert the secondary inside the primary after having wound heavy wrapping paper around the primary until it fits closely into the larger tubing. The secondary support with its brass rods firmly secured to it is then to be brought up to the wooden head of the secondary and holes drilled for the wood screws that secure the secondary to the support. The shrewd reader will at once see that this insures accuracy in assembling the component parts of the transformer and precludes the possibility of uneven space between primary and secondary cylinders. When the screws have been driven home, the secondary may be withdrawn and from this point on the secondary and its supporting piece should not be separated.

The next operation will be to lay out the arc of the circle for the contact points, of which there are 13. The radius of the arc on

strument. The details and dimensions are given in Fig. 3. The choice of wood rests with the individual, but we favor whitewood. The base, 7½ in. wide and 18 in. long, is of ⅞ in. stock and quite simple. The upright pieces that support the primary are each 6½ in. wide and 7 in. high, while the thickness may be ⅞ in. The piece to the right in Fig. 3 is cut out to receive the cardboard cylinder which is 5½ in. outside diameter. The left-hand piece is left solid and it has mounted upon it a disc of ⅜ in. whitewood turned to fit the inside of the cylinder. The latter should not be permanently secured in its supports until after the holes for the brass tubes indicated in the drawing have been drilled. This operation will be detailed later.

The next step will be to get out the two discs for the secondary cylinder. One of these is of ⅜ in. stock, while the other is of ⅞ in. wood. They are, of course, to fit the inside of the cylinder tightly. The latter, as the drawing indicates, is 5 in. in outside diameter. The internal diameters of both cylinders will vary a trifle with different makes of cardboard tubing, but it is of little consequence. The thicker disc may be permanently affixed by means of glue and wooden pegs, but the thinner disc must be left removable until after the winding is finished and connections to the contact points are made.

The front support that holds the secondary may now be worked out with the jig saw, and with this the woodwork proper is finished. This is the time to do the staining and varnishing. If these decorative operations are left until later, the results will be rather unsatisfactory and the task difficult to perform. After the varnish is good and hard, so that handling will not injure the finish, the worker may lay out the centers for drilling the two holes through which pass the supporting brass rods. These holes are 4⅜ in. apart on centers and 1¼ in. up from the base. When the centers have been marked and pricked, the front support may be placed on two pieces which hold the primary, and holes drilled through all three while they are clamped together. This will insure alignment of the holes, which is quite essential to prevent binding when the rods are telescoped into the tubes. The holes in the secondary support may be ¼ in., the same as those in the other pieces, for a trifle of leeway is not objectionable to permit of easing up on any bind that may develop when the apparatus is assembled.

The brass tubing and the rods that enter it may then be placed, and the whole arrangement assembled temporarily to determine whether the work has been successful. As the drawing indi

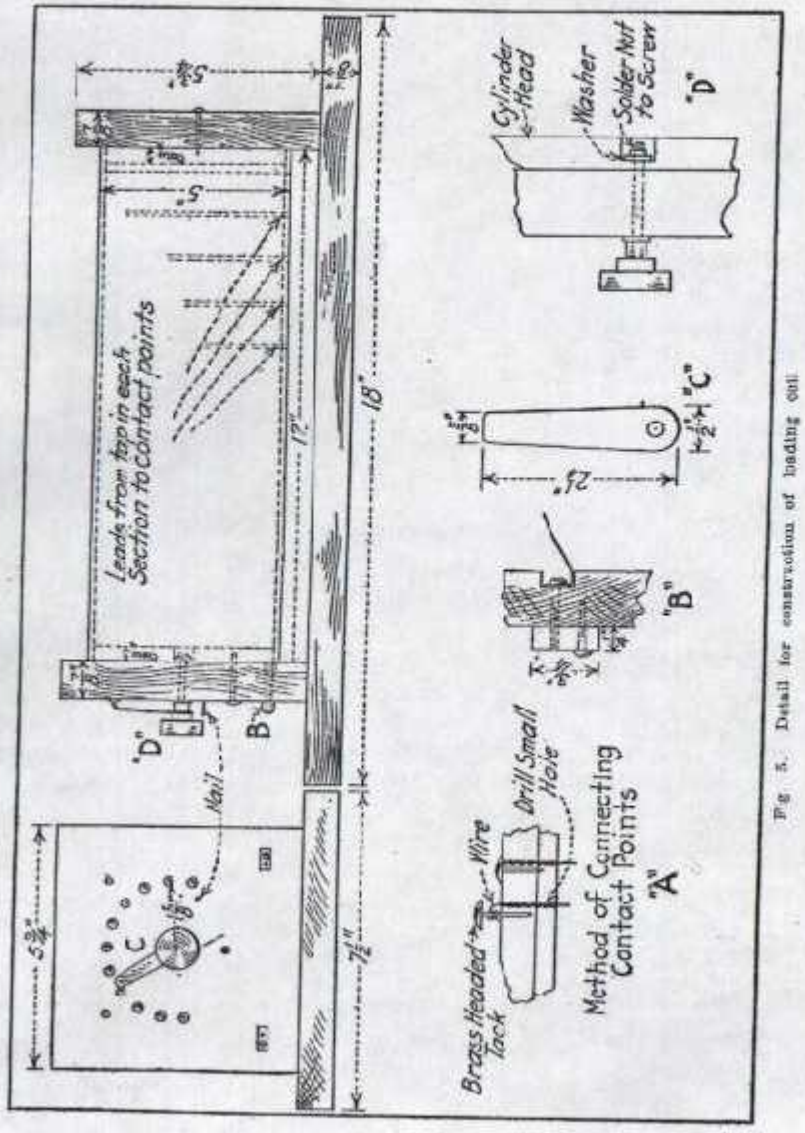


Fig 5. Detail for construction of loading coil

shape, and soldered to the bottom of the slider.

Binding posts are not specified, for connection may be made directly to machine screws in the ends of the slider rods in the case of the primary, and to small connection blocks electrically connected with the brass tubes that pass through the primary supports, for the secondary.

Loading Inductance

The loading inductance consists of a cardboard cylinder wound in a single layer with No. 24 S. S. C. copper magnet wire in 11 sections. Each section is tapped to a contact point in one of the heads as shown in the illustration. As the construction is identical with that of the receiving transformer just described, in so far as mounting the cardboard cylinder, etc., is concerned, we will not go into a lengthy explanation of the construction. All details are given very clearly in Fig. 5, and Fig. 4 shows well the appearance of the finished instrument.

A few hints relative to the standard products used in the construction of both of these coils may not be amiss. The instrument switch used on each coil is a standard product that may be bought in almost any large supply house. The builder is, however, advised to discard the usual brass contact piece and substitute for it a piece of phosphor bronze sheeting cut to the dimensions shown at *C*, Fig 5, and bent over at the tip so that the contact is made on the edge rather than on the flat surface.

The contact points may be ordinary brass-capped upholstery tacks or they may be oval-head copper rivets. The latter are much to be preferred, but they are more expensive and more difficult to install. If the rivets are used, however, the hole through which the tap wire passes may be made approximately the same size as the shank of the rivet in order that the latter may make contact as the stud is driven home.

If the upholstery tacks are used, the method of connecting shown at *A* should be employed. This insures contact between the wire and the brass head rather than with the steel shank of the tack.

The simple form of connection block that we favor is shown at *B*. This is merely a short length of ¼ in. square brass rod secured to the instrument at the desired point. Connection is made under the heads of brass machine screws in an obvious manner.

The cardboard cylinders for both receiving transformer and loading coil may be obtained from advertisers in this magazine, and the woodwork is best obtained from a local mill which will cut out the pieces to size at a very reasonable figure.

At *D* is shown the method we employ to secure the adjusting

knob which carries the switch making contact with the studs. In the standard product a threaded shank is permanently fastened to the composition knob. The obvious way of securing this switch to the instrument would be merely to place a lock-nut over the first nut that goes on the shank. This method, however, is unreliable and unsatisfactory. Constant use will frequently loosen the nuts after the instrument is entirely assembled, and needless annoyance results therefrom. We used the simple expedient of locking the first nut with a drop of solder after we had secured just the tension we desired to make the knob turn with freedom but without unpleasant looseness.

The *Technical Adviser* is at the disposal of readers who require additional advice or instructions, and the *Service Department* will aid those who have difficulty in obtaining needed materials.

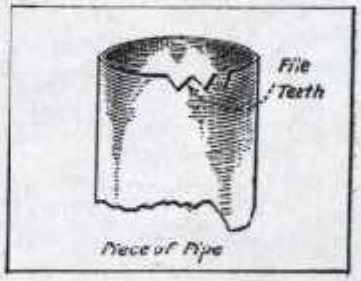
Paint For One Cent Per Pound

To one gallon of hot, soft water add four pounds of crude sulphate of zinc. Let it stand until it dissolves perfectly, and a sediment will settle at the bottom. Turn the clear solution into another vessel. To one gallon of paint (lead and oil) mix one gallon of the compound. Stir into the paint slowly for 10 or 15 minutes and the paint and compound will combine perfectly. If too thick, thin the mixture with turpentine.

Contributed by J. C. Gilliland.

Boring a Hole in Brick

Any man who wishes to bore a hole in brick and has not a cold chisel on hand will find this home-made chisel very handy. All that it consists of is a piece of galvanized iron pipe about 7 in. long and the diameter de-



File teeth in one end of the pipe

pends upon the hole you want to bore. In one end of this pipe teeth must be filed about 1-16 in. apart. You will find that this will bore a clean-cut hole.

Contributed by Fred W. Allen.

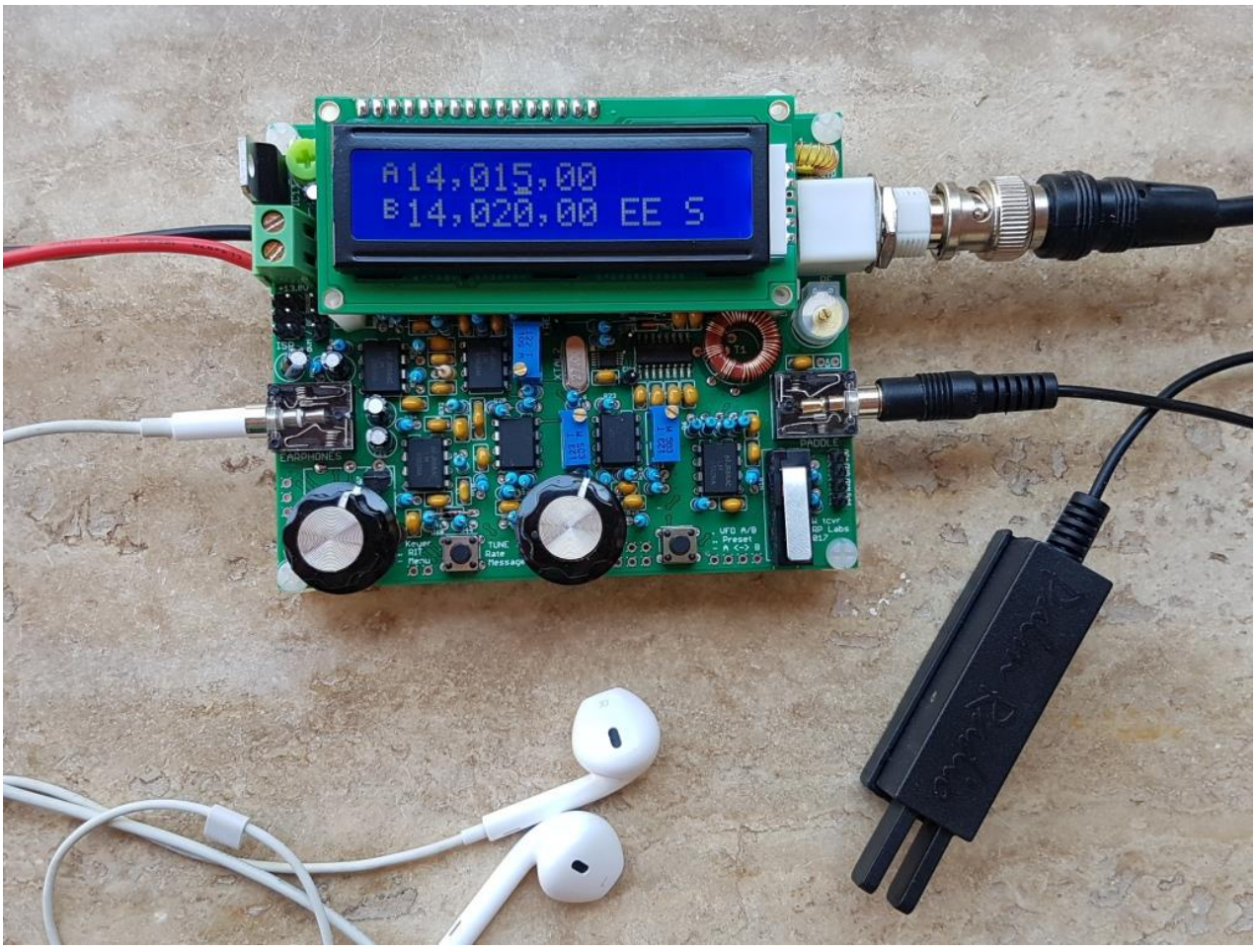
If the family is large and the kitchen sink small, try using an oval tin foot tub instead of the round dish pan.—Mary F. Scott.

QCX CW Transceiver

QCX 5W CW Transceiver kit assembly instructions

The “QCX”: a single band, high performance 5W CW Transceiver with built-in alignment and test equipment, iambic keyer, WSPR beacon mode, and more...

Designed and produced by QRP Labs, 2017



(pictured with Palm Radio pico paddle http://palm-radio.de/english/eppaddle.html**)**

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1. Introduction

Thank you for purchasing this high performance single-band 5W CW transceiver kit, the QCX (for **Q**RP Labs **C**W **X**cvr). This kit has a long list of features!

- Easy to build, single-board design, 10 x 8cm, all controls are board-mounted
- Professional quality double-sided, through-hole plated, silk-screen printed PCB
- Choice of single band, 80, 60, 40, 30, 20 or 17m
- Approximately 3-5W CW output (depending on supply voltage)
- 7-16V recommended supply voltage
- Class E power amplifier, transistors run cool… even with no heatsinks
- 7-element Low Pass Filter ensures regulatory compliance
- CW envelope shaping removes key clicks
- High performance receiver with at least 50dB of unwanted sideband cancellation
- 200Hz CW filter with no ringing
- Si5351A Synthesized VFO with rotary encoder tuning
- 16 x 2 blue backlight LCD screen
- Iambic keyer or straight key option included in the firmware
- Simple Digital Signal Processing assisted CW decoder, displayed real-time on-screen
- On-screen S-meter
- Full or semi QSK operation using fast solid-state transmit/receive switching
- Frequency presets, VFO A/B Split operation, RIT, configurable CW Offset
- Configurable sidetone frequency and volume
- Connectors: Power, 3.5mm keyer jack, 3.5mm stereo earphone jack, BNC RF output
- Onboard microswitch can be used as a simple straight Morse key
- Built-in test signal generator and alignment tools to complete simple set-up adjustments
- Built-in test equipment: voltmeter, RF power meter, frequency counter, signal generator
- Beacon mode, supporting automatic CW or WSPR operation
- GPS interface for reference frequency calibration and time-keeping (for WSPR beacon)

No test equipment is required to build, align and operate this CW transceiver. Its innovative self-alignment and self-test features will help and guide you in setting up the transceiver in a few easy steps. The kit also includes a voltmeter, RF power meter, frequency counter and signal generator which can aid in debugging and fault-finding.

We hope you enjoy building and operating this kit! Please read this assembly manual carefully, and follow the instructions step by step in the recommended order. Later in the manual the circuit design is described in detail and we recommend reading and understanding this section too, to get the maximum enjoyment and education from your new radio.

Typical performance measurements are shown in the measurements section. The operation section of the manual describes transceiver, alignment and test equipment operation in detail. **There is a single page reference "cheat sheet" near the end of the manual.**

2. Parts list

Resistors (all fixed resistors are ¼-Watt, 1% tolerance)

R5, R6, R8, R9	100-ohm (4 pcs), brown-black-black-black-brown
R48	270-ohm, red-purple-black-black-brown
R41	470-ohm, yellow-purple-black-black-brown
R3, 4, 19, 26, 37, 42, 45, 53-55, 62, 63	1K (12 pcs), brown-black-black-brown-brown
R12, 13, 15, 16, 20, 22, 23, 25, 44, 56, 59	3.3K (11 pcs), orange-orange-black-brown-brown
R61	3.9K, orange-white-black-brown-brown
R18	4.3K, yellow-orange-black-brown-brown
R11	5.1K, green-brown-black-brown-brown
R21	7.5K, purple-green-black-brown-brown
R1, 2, 7, 10, 14, 34, 39, 40, 46, 49-52, 57, 58	10K (15 pcs), brown-black-black-red-brown
R28, 29	33K (2 pcs), orange-orange-black-red-brown
R30, 31	47K (2 pcs), yellow-purple-black-red-brown
R32, 33	36K (2 pcs), orange-blue-black-red-brown
R38, 43, 60	120K (3 pcs), brown-red-black-orange-brown
R35	750K, purple-green-black-orange-brown
R27	500-ohm multi-turn trimmer potentiometer
R17, 24	50K multi-turn trimmer potentiometer (2 pcs)
R36	5K Logarithmic potentiometer
R47	100K trimmer potentiometer

Capacitors

C5	22pF, “220” (not used on every band version)
C30	30pF, “300” (not used on every band version)
C8	39pF, “390” (not used on every band version)
C14, 16, 18, 23, 33	1nF, “102”, (5 pcs)
C19, 20	2.2nF, “222” (2 pcs)
C53	3.3nF, “332’
C4, 7, 10, 42	10nF, “103” (4 pcs)
C15	33nF, “333”
C17	39nF, “393”
C9, 13	47nF, “473” (2 pcs)
C2, 3, 6, 12, 29, 32, 34-36, 39-41, 48-50, 52	0.1uF, “104” (16 pcs)
C11, 43-46	470nF, “474” (5 pcs)
C31	1uF, “105”
C21, 22, 24, 37, 38, 51	10uF electrolytic (6 pcs)
C47	470uF electrolytic
C25, C26	Value depends on band, in LPF kit bag (2 pcs)
C27, C28	Value depends on band, in LPF kit bag (2 pcs)
C1	30pF trimmer

Semiconductors

D1, 2, 4, 5	1N4148 (4 pcs)
D3	1N5819

IC1	SI5351A 10-pin MSOP (pre-soldered at factory)
IC2	ATmega328P 28-pin microcontroller
IC3	74ACT00N 14-pin quad NAND gate
IC4	FST3253 16-pin SOIC (pre-soldered at factory)
IC5-10	LM4562 8-pin dual op-amp (6 pcs)
IC11	7805, TO220 5V 1A voltage regulator
Q1-5, 7	BS170 TO92 MOSFET (6 pcs)
Q6	MPS751 TO92 transistor (MPS2907 in early kits)

Inductors

L1, 3	Value depends on band, in LPF kit bag (2 pcs)
L2	Value depends on band, in LPF kit bag
L4	T37-2 (red) toroid, value depends on band
L5, 6	100uH axial molded inductors (2 pcs)
T1	T50-2 (red) toroid, value depends on band

Miscellaneous

2 x 3-pin	Male header connector for in-circuit programming
4-pin	Male header connector, in LPF kit bag (2 pcs)
16-pin	Male header connector
16-pin	Female header connector
2-pin terminal block	2-pin screw terminal block
3.5mm stereo	3.5mm stereo connector (2pcs)
BNC	PCB-mounting BNC connector
S1	Microswitch
S2, 3	6 x 6mm push-to-make button (2 pcs)
SW1	Rotary encoder with shaft push-button
16 x 2 LCD	HD44780 LCD 1602 with blue backlight
XTAL1	20MHz HC49/4H quartz crystal
XTAL2	27MHz HC49/4H quartz crystal
PCB	4.0 x 3.2-inch PCB (101.6 x 81.3mm)
spacer	12mm M3 nylon hex spacers (8 pcs)
screw	6mm M3 nylon screws (12 pcs)
Knob	Knobs to fit rotary encoder and R36 (2 pcs)

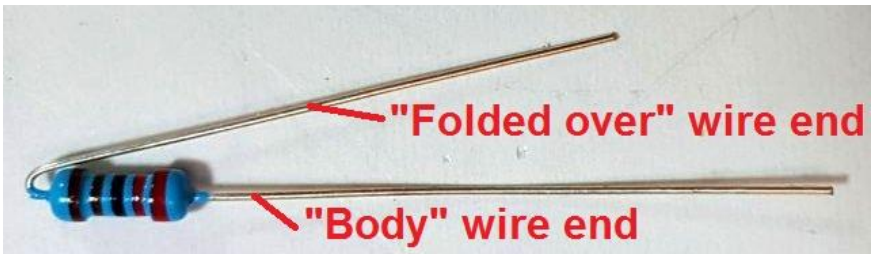
3. Assembly

Assembly of this kit is quite straightforward. But there are quite a lot of components. So please keep them methodically in trays or some convenient storage boxes, and be careful not to misplace any. The usual kit-building recommendations apply: work in a well-lit area, with peace and quiet to concentrate. **The IC (chips) and some of the other semiconductors in the kit are sensitive to static discharge. Therefore, observe Electrostatic discharge (ESD) precautions**. And I say it again: **FOLLOW THE INSTRUCTIONS!!** Don't try to be a hero and do it without instructions!

A jeweller's loupe is really useful for inspecting small components and soldered joints. You'll need a fine-tipped soldering iron too. It is good to get into the habit of inspecting every joint with the magnifying glass or jeweller's loupe (like this one I use), right after soldering. This way you can easily identify any dry joints or solder bridges, before they become a problem later on when you are trying to test the project.

Triple check every component value and location BEFORE soldering the component! It is easy to put component leads into the wrong holes, so check, check and check again! It is difficult to de-solder and replace components, so it is much better to get them correctly installed the first time. In the event of a mistake, it is always best to detect and correct any errors as early as possible (immediately after soldering the incorrect component). Again, a reminder: removing a component and re-installing it later is often very difficult!

All of the resistors in the kit are installed vertically, because PCB space is limited. You will need to bend the resistor wires as shown in the picture below. On the parts layout diagram, there is a small circle around the hole where the body end of the resistor should be installed. This is not critical (resistors don't care which way around they are) but it is recommended, just to try to ensure that the folded over wire-end of the resistor doesn't touch any other wire or metal part.



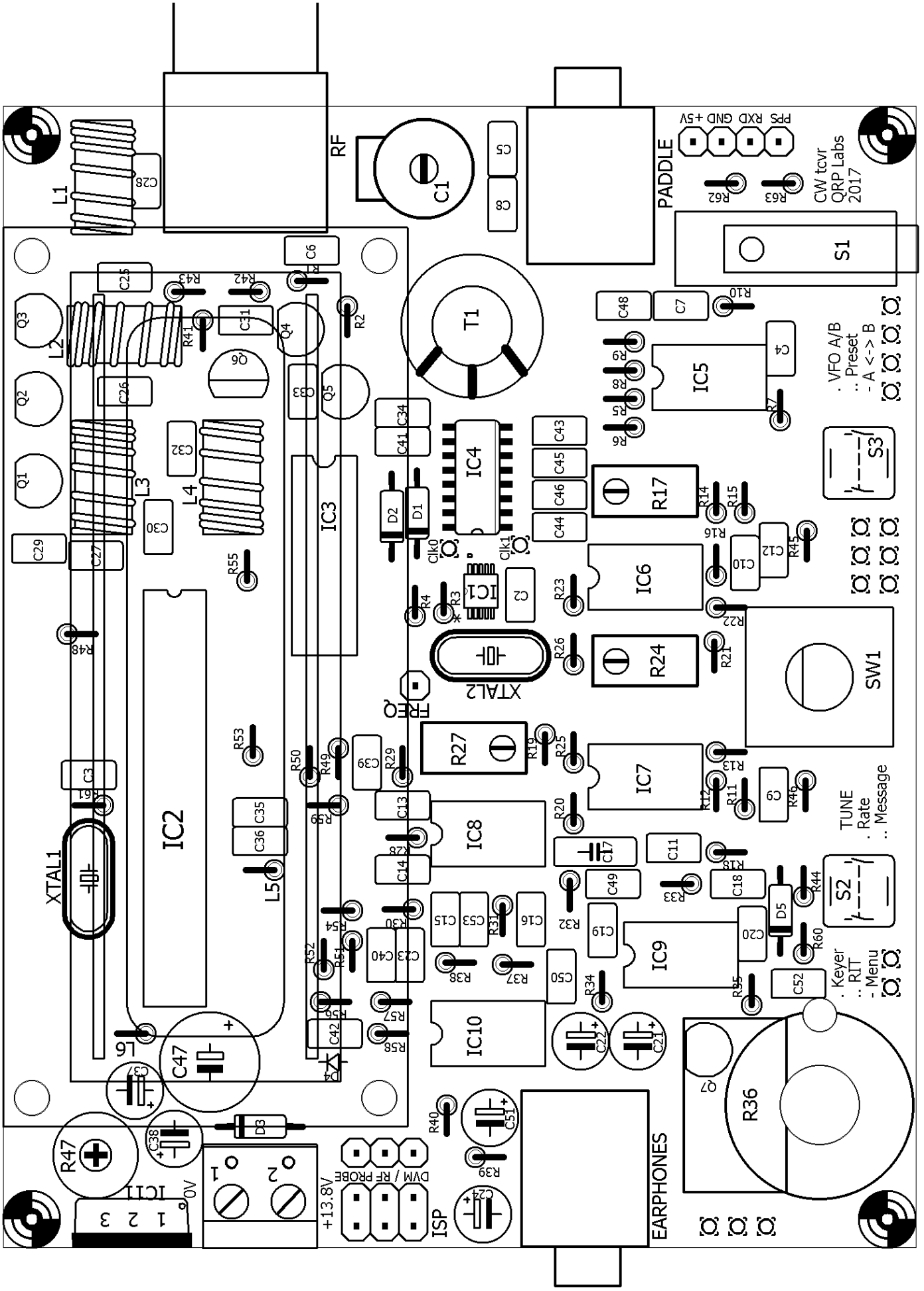
Please refer to the layout diagram and PCB tracks diagrams below, and follow the steps carefully.

Assembly steps will be in the order of smallest to largest components. I follow the order semiconductors, capacitors, resistors and finally all other (generally larger) components. It is probably unnecessarily thorough and complex to build the radio one stage at a time and test each stage one by one… I recommend just install everything then power up.

As per standard QRP Labs practice, the ATmega328P microcontroller has a 28-pin DIP socket in case you may wish to subsequently replace it for firmware upgrades etc. The Si5351A and FST3253 ICs are only available in surface mount packages so these are already soldered to the PCB for you, at the factory. All other components used are all leaded through-hole packages, and all are installed on the top side of the PCB.

Some people like using IC sockets, some don't. Undoubtedly there is a cost impact, of course. Personally, I believe that on average, the use of IC sockets causes more problems than it solves. They create the potential for bent pins that don't make contact properly, hidden shorts under the socket, issues with increased lead length, etc. Accordingly, the remaining ICs are soldered directly into the PCB without IC sockets. Please forgive me if you disagree.

You may find it convenient to inventory the resistors and capacitors and sort them by value, so that you can easily find the correct ones as you work through the assembly steps.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PPS RXD GND +5V
RF
EARPHONES
ISP
DVM / RF PROBE
+13.8V
. VFO A/B
.. Preset
- A <-> B
TUNE
. Rate
.. Message
. Keyer
.. RIT
- Menu
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IC10
IC11
S1
S2
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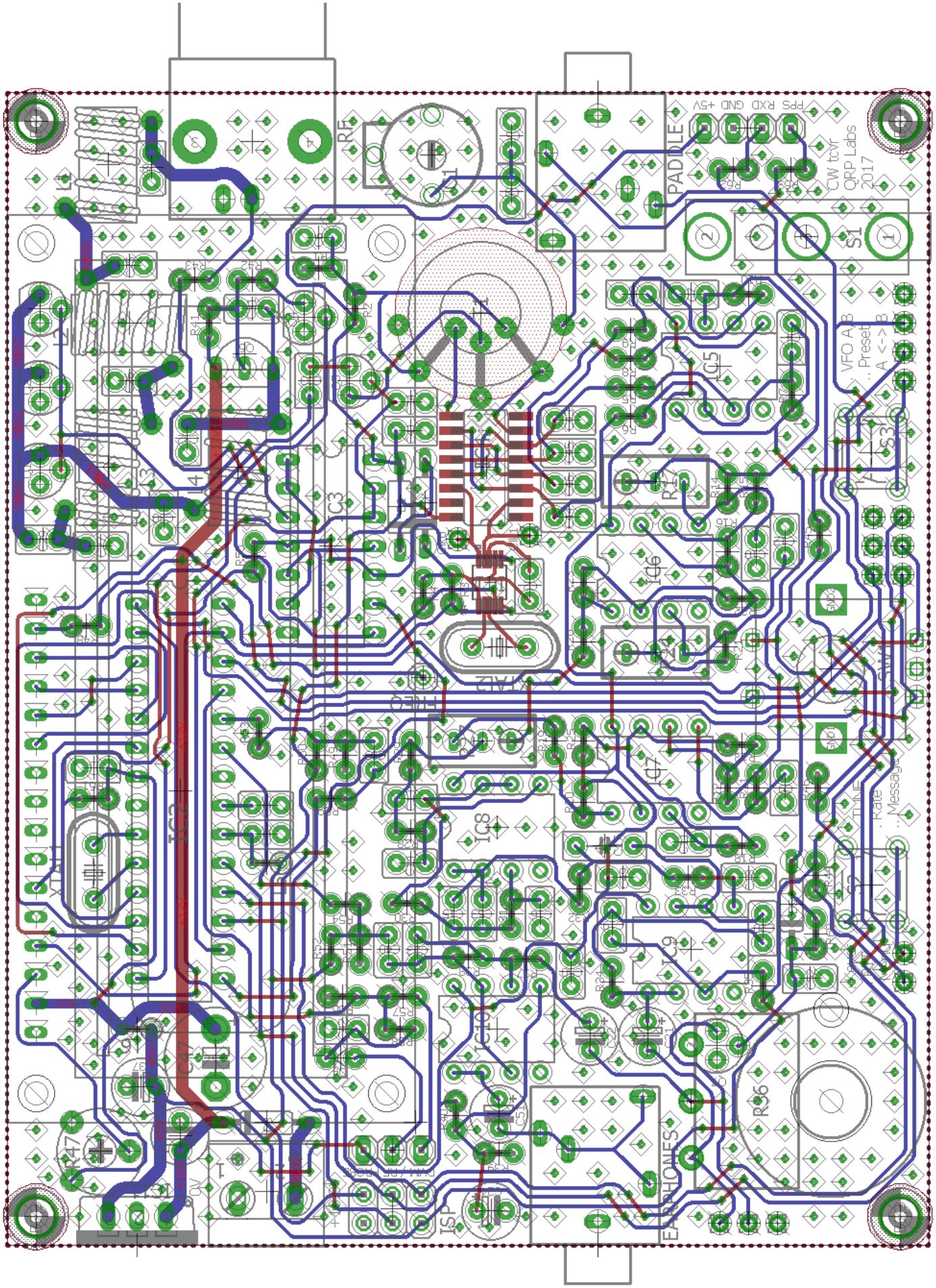
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The PCB track and circuit (schematic) diagrams are shown on the previous pages. Tracks shown in BLUE are on the bottom layer. Tracks shown in RED are on the top layer. There are only two layers (nothing is hidden in the middle). Not shown in this diagram are the extensive ground-planes, on both sides of the board. Practically everything on both layers that isn't a RED or BLUE track, is ground-plane! The two ground-planes are connected at frequent intervals (not more than 0.1-inches) by vias. This is the kind of layout I have done previously for a quad-band GSM device operating at up to 1900MHz… it is probably overkill in an HF transceiver… but if you can, why not! I used to say often that you can never have too much supply line filtering and decoupling, and never have too much shielding. Both these statements don't apply so conveniently to kits as they do to homebrew projects. In a kit every decoupling capacitor has a cost in both money and PCB area (which also means more money). Shielding is even more difficult and expensive. So shielding and decoupling should be applied where needed only! But ground-plane – well that's another story. It's free, and without drawbacks – so why not, let's just put it everywhere.

The PCB silkscreen component numbers are necessarily quite small and may be overlapping pads or vias, or near to other components. It may be nice to have doubled the size of the PCB and have more space to print nice silkscreen labels of the component numbers. But the result wouldn't be such a cute and neat little radio and the price wouldn't be so friendly either! So check carefully, before soldering each component! Check the location and check again – with reference to the diagrams in each step.

All components are installed on the top (component side) of the PCB and soldered on the bottom (solder side) of the PCB.

Take care when installing integrated circuits. All through-hole integrated circuits are supplied by the manufacturers with their pins bent a little wide. You need to carefully bend the rows of pins together a little, in order to fit them in the PCB holes, or in the 28-pin IC socket in the case of the ATmega328P microcontroller.

The band-specific Low Pass Filter (LPF) parts are supplied in a separate LPF kit bag.

In the construction for some bands, not all of the capacitors supplied in the kit are used. Do not be alarmed if you have a few components left over at the end!

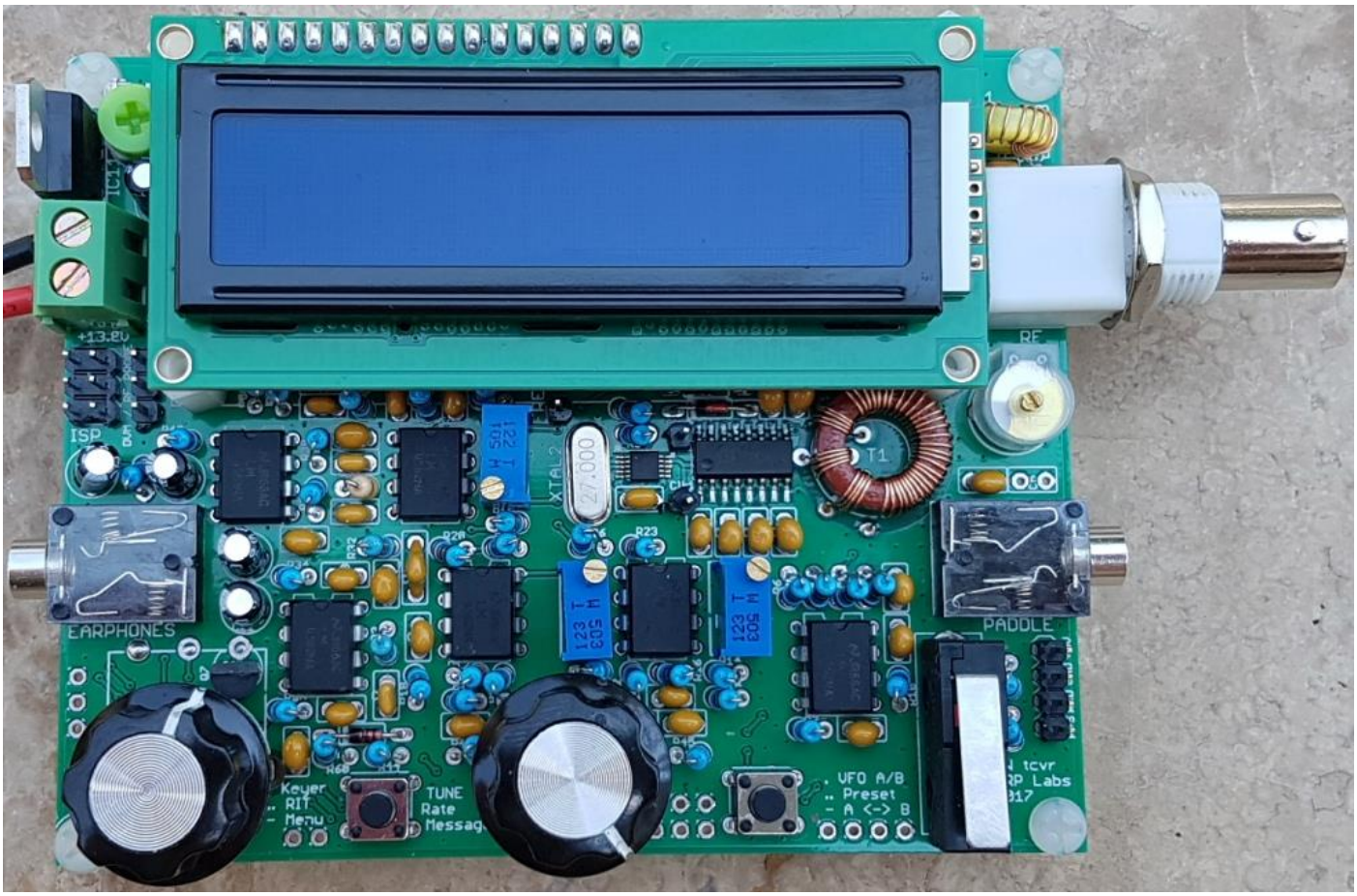
Wind the L1-3 inductors with the enamelled copper wire supplied in the LPF kit bag. Wind the other inductors (L4 and transformer T1) using the wire supplied in the main kit bag.

The component colour coding of the layout diagram at every step of the assembly instructions is as follows (kind of: components past, present and future):

- Components shaded grey have already been installed
- Components shaded red are the ones being installed in the current assembly step
- Components shaded white are the ones which have not yet been installed

The following photographs show the final assembly, with and without the LCD module plugged in. You can keep these photographs in mind when assembling the kit, they will give you some idea of how the kit fits together and help avoid assembly errors.

XTAL1
20.000
L1
RF
T1
XTAL2
27.000
ISP
EARPHONES
PADDLE
+13.8V
Keyer
RIT
Menu
TUNE
Rate
VFO A/B
Preset
A <-> B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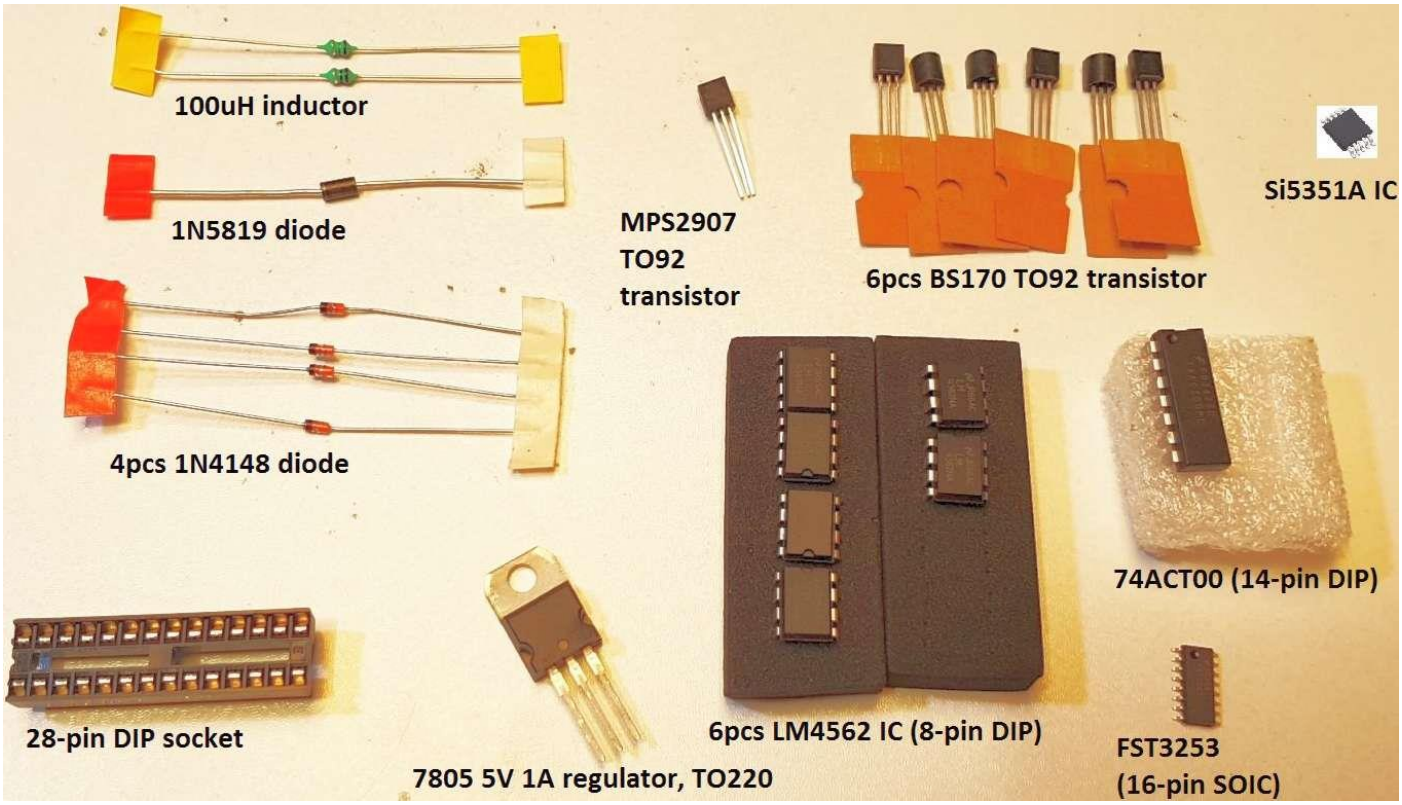
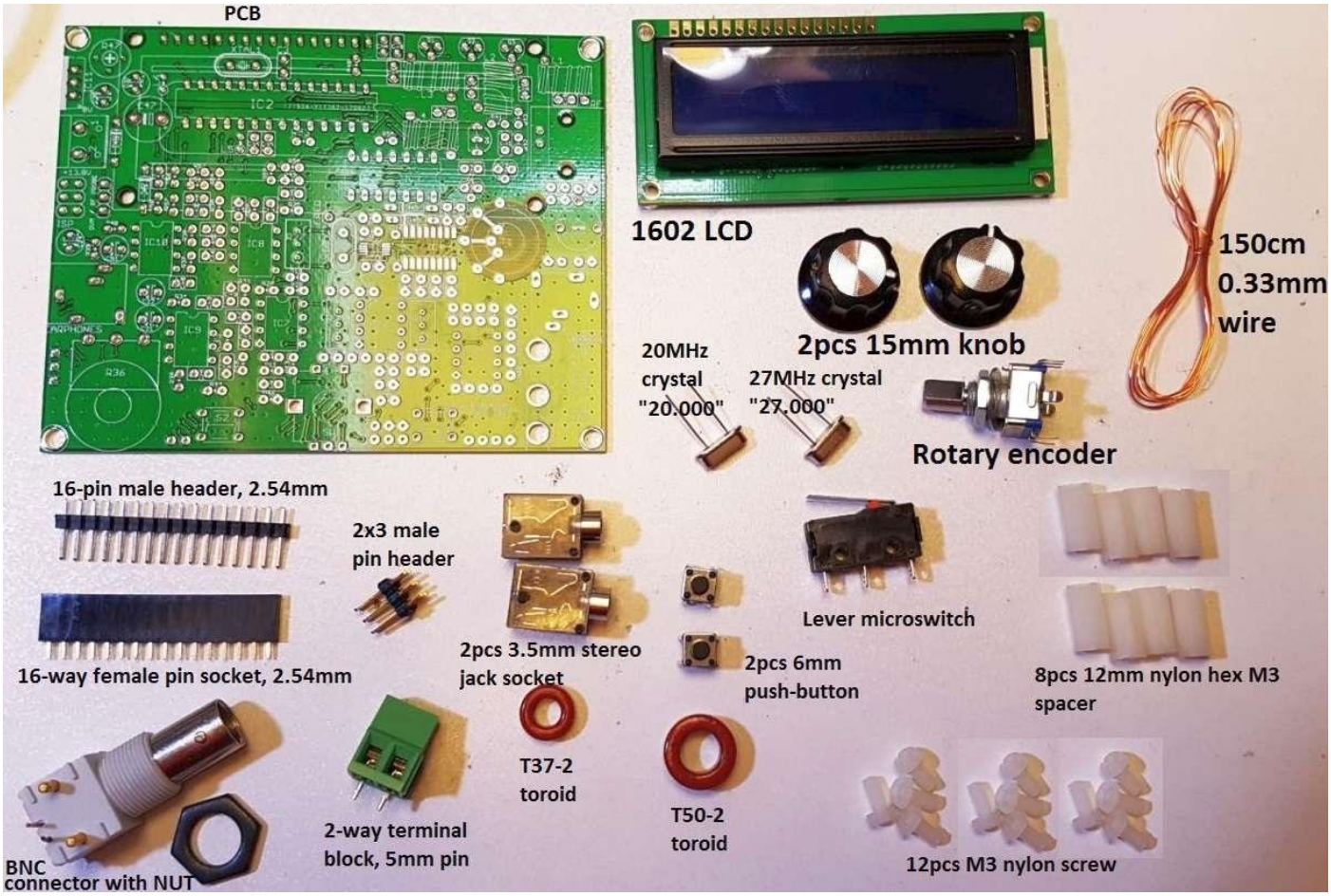
3.1 Inventory parts

Refer to parts list in section 2. The following photographs are to aid component identification. The resistors and capacitors are omitted.

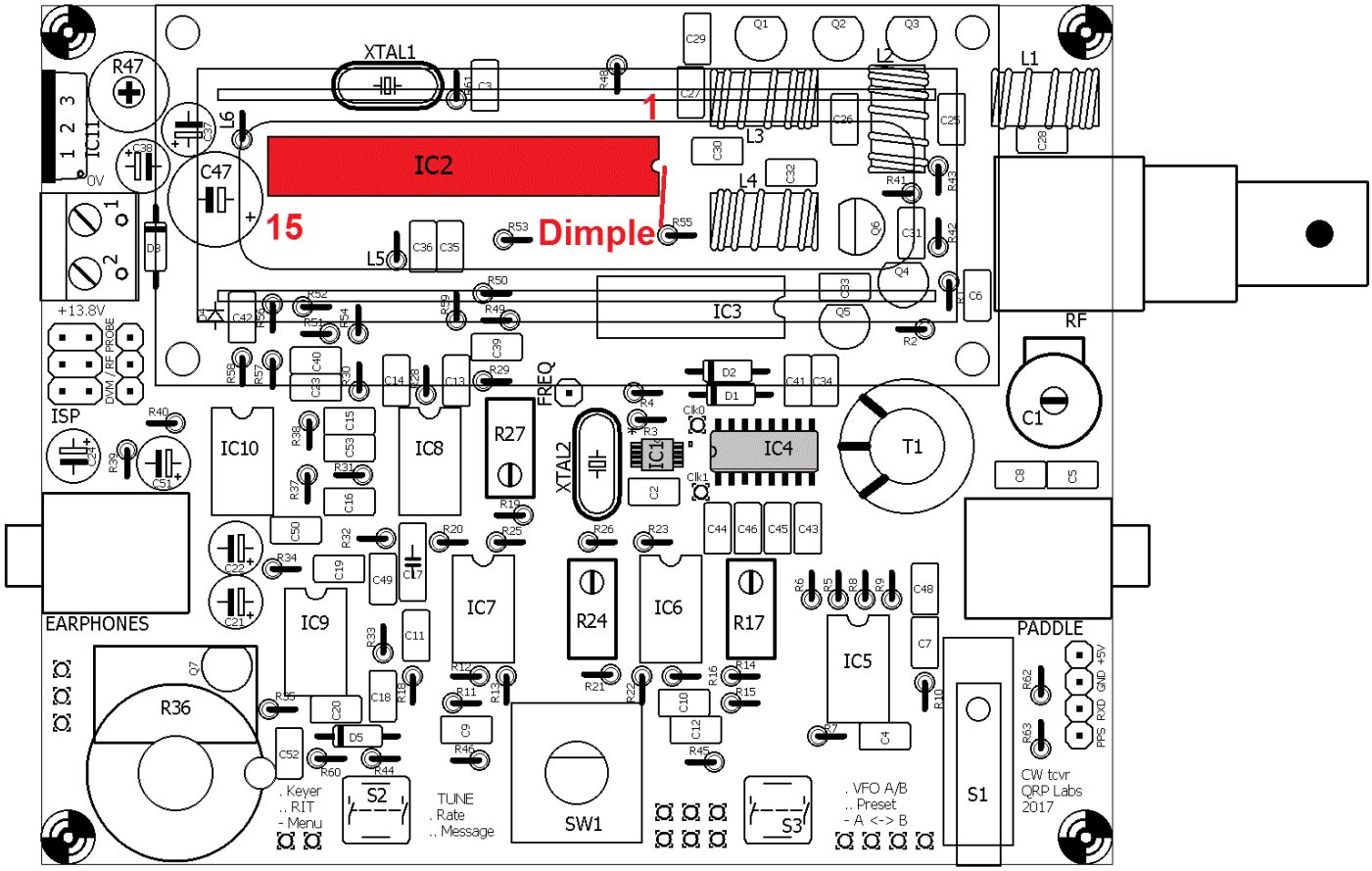




3.2 Install IC2 socket

Install the 28-pin IC socket for IC2. Take care to match the dimple on the socket, with the dimple on the PCB silkscreen. It is critical to insert the microcontroller with the correct orientation. Lining up the dimple on the PCB silkscreen, socket and actual IC is the best way to avoid confusion and potential error.

I recommend soldering one pin at each diagonal, for example pins 1 and 15. You can then check that the IC socket is seated firmly on the PCB, and correct any issues easily. Once all the pins are soldered it will be difficult, if not impossible, to change anything. When you are happy with the position of the socket, proceed with soldering the other 26 pins.



3.3 Install DIP integrated circuits IC3, IC5-10

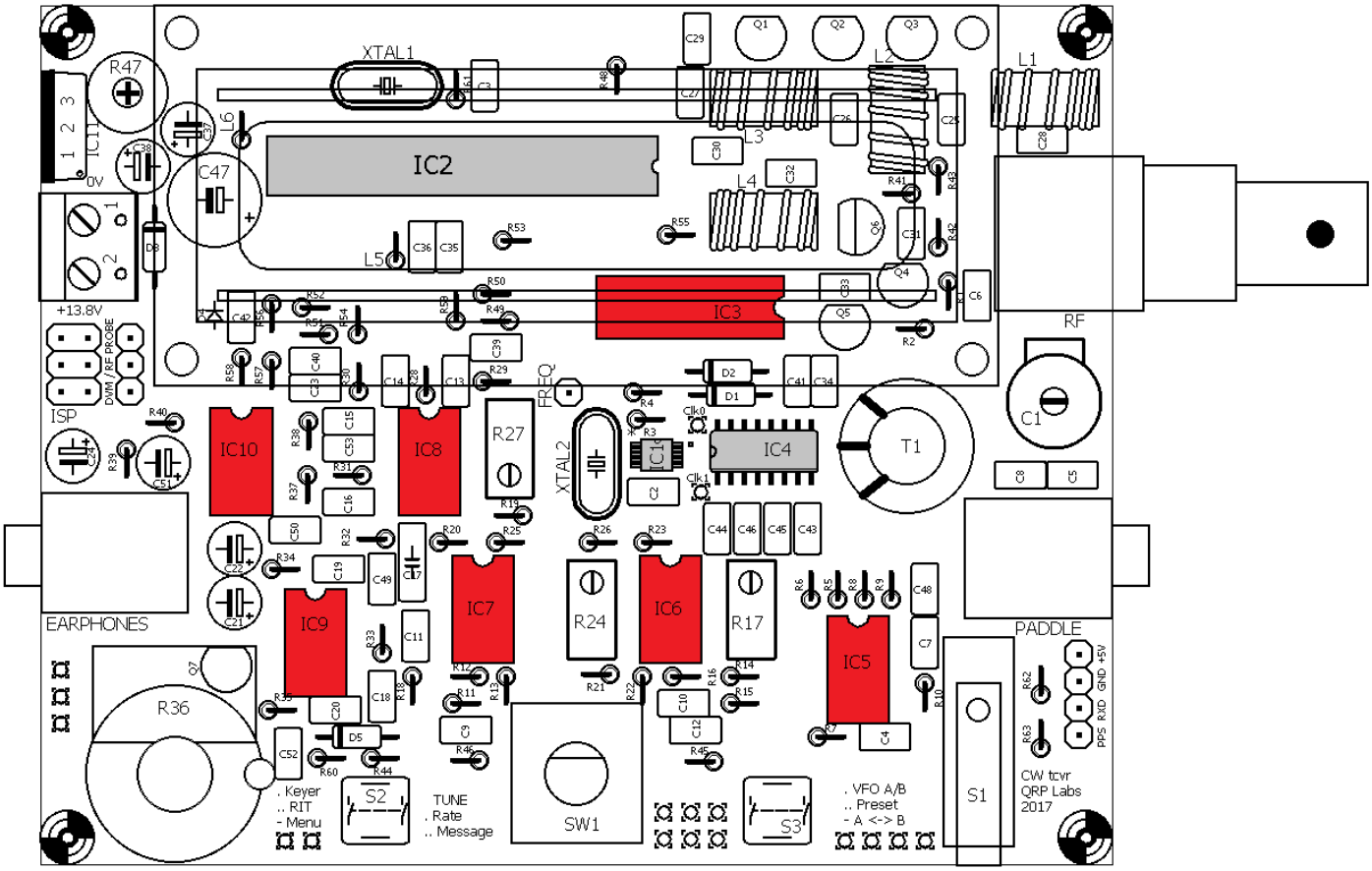
In the same way, install the Dual-Inline-Package integrated circuits IC3 and IC5-10. We install these now because they have more pins than all the passive components (resistors, capacitors etc) and it is easiest to reach and install the ICs properly when they are alone on the board. All of the operational amplifier ICs (IC5-IC10) are part number LM4562, so it is impossible to mix them up, they are the only 8-pin DIP chips in the kit. Similarly, IC3 is the only 14-pin DIP chip.

Electrostatic Discharge (ESD) precautions should be observed though personally I do not think it is necessary to get crazily paranoid about this. Don't wear 100% nylon clothing and dance around on a nylon carpet etc to build up static charge! But I don't go to the other extreme of earth mats, earthed wrist bands etc.

It IS absolutely critical to align the IC's properly! Be very careful to check and double-check and even triple-check that the dimple in each IC package is aligned with the dimple in the silkscreen print on the PCB (and the layout diagram).

It is normal to need to slightly squeeze the rows of pins inwards to fit in the holes on the PCB. IC manufacturers always supply leaded (through-hole) ICs with the pins a little too wide-spaced.

As before, I find it useful to solder initially two pins at diagonal opposites of the chip, then check that the chip is nicely seated evenly on the board, before soldering the other pins.

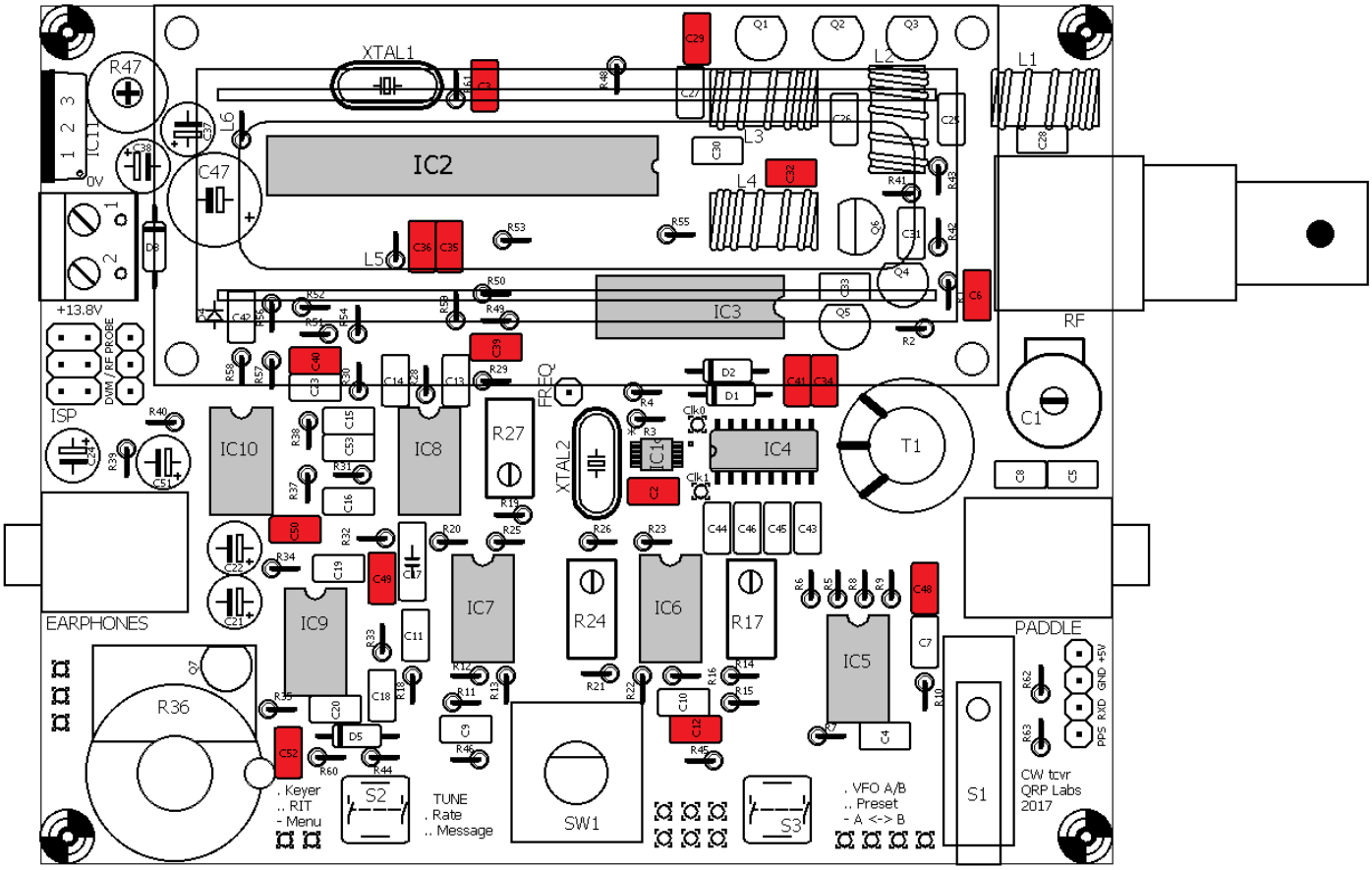


3.4 Install all 100nF (0.1uF, “104”) capacitors

There are 16 100nF (0.1uF) capacitors, these have the code “104” written on them. Be sure to identify the correct capacitors, using a magnifying glass or jeweller’s loupe. These capacitors are C2, C3, C6, C12, C29, C32, C34, C35, C36, C39, C40, C41, C48, C49, C50 and C52. Place each in the correct position on the board, and slightly bend the legs outwards at about 30-degrees angle so that they stay in place. Then solder the wires, and trim the excess wire length with wire cutters.

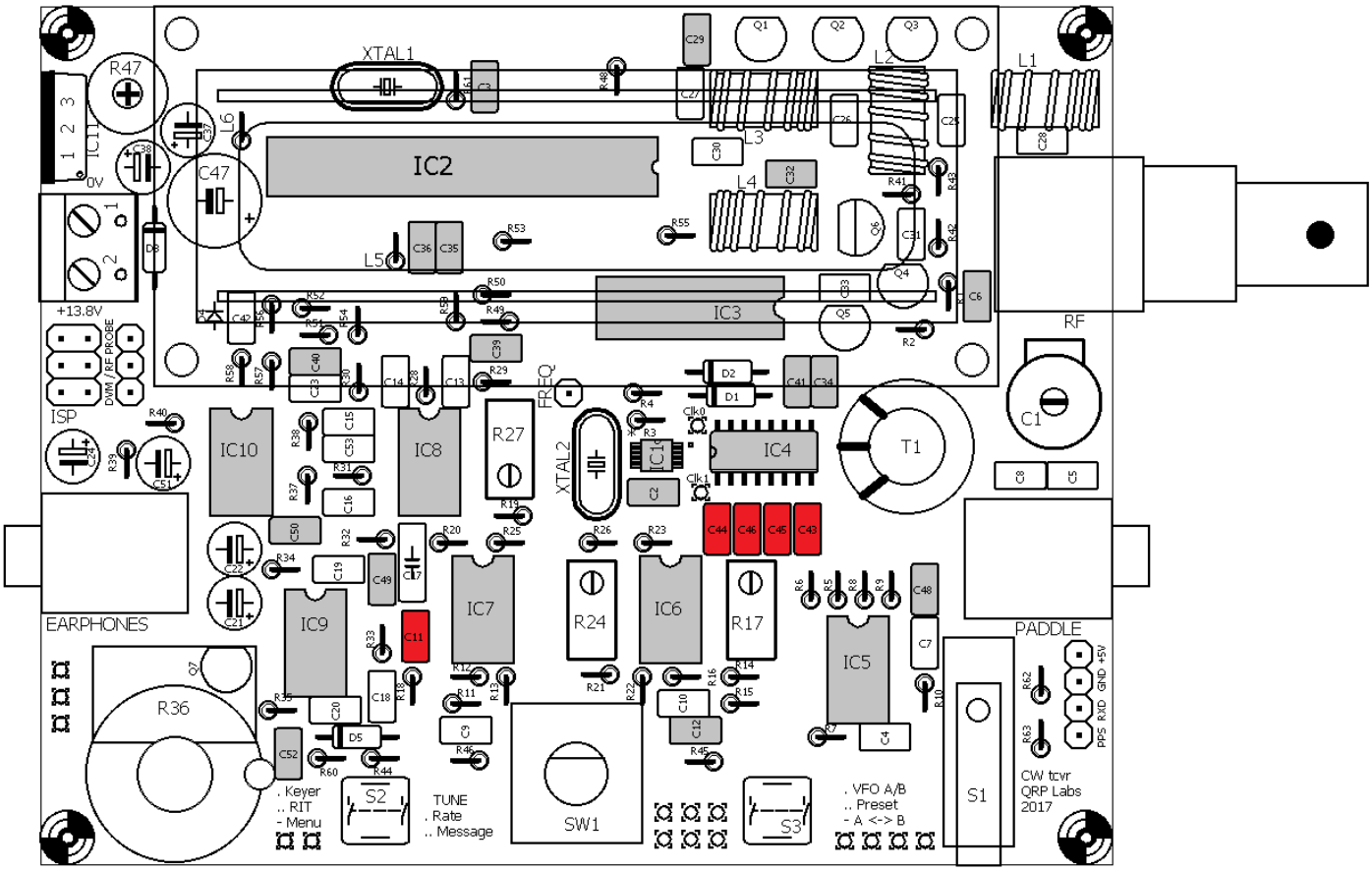
It does not matter which way round the capacitors are installed. However, it is very good practice to install them all with the capacitor label facing in the same direction. For example, ensure all the “104” labels face to the front, or to the right (depending which way the capacitor is orientated on the PCB). This makes it much easier to inspect the PCB assembly later.

Take particular care soldering wires which connect to the ground plane. Despite the “thermals” (a ground pad is connected to the ground plane by four thin traces, not a continuous ground plane, to make soldering easer), the heat dissipation is still more, and it can be harder to make a good joint.



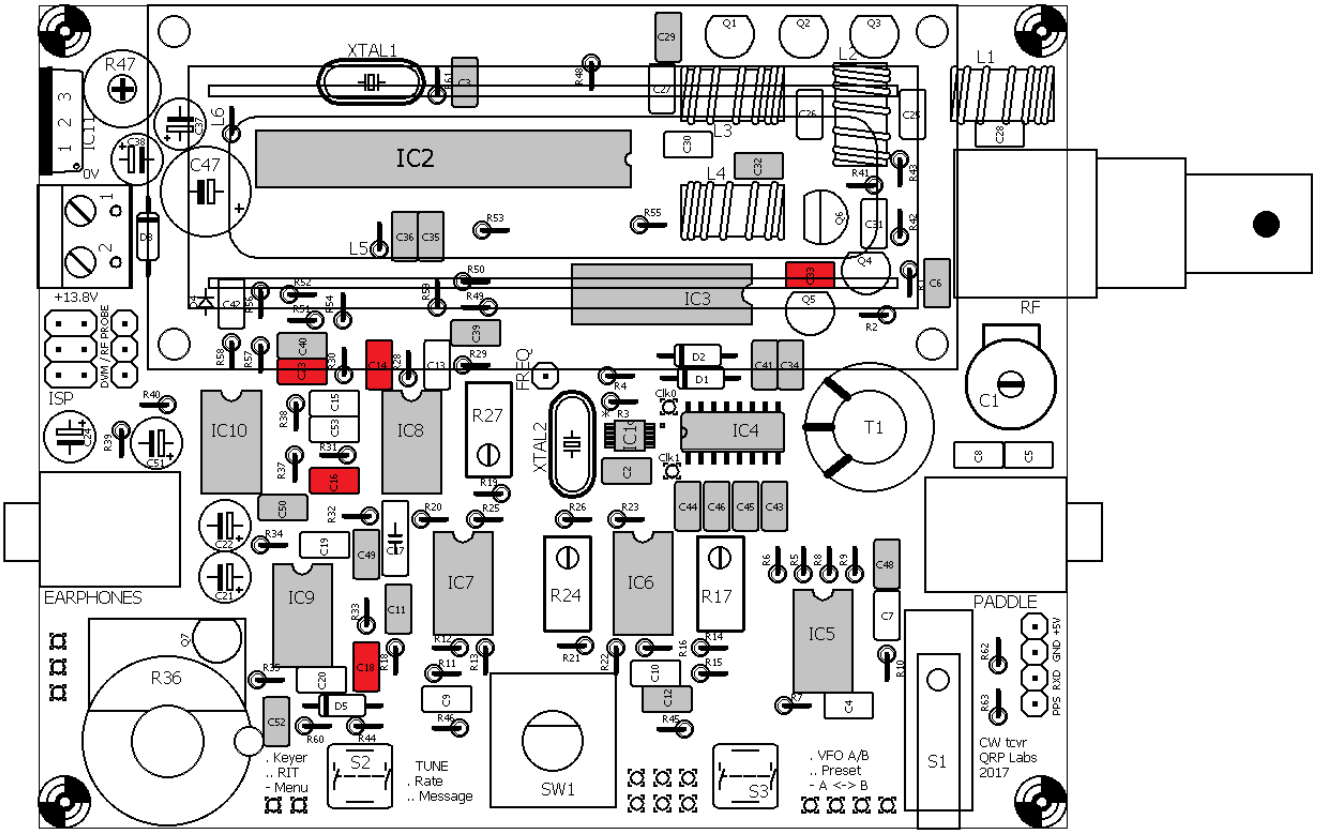
3.5 Install all 470nF, “474” capacitors

The 470nF capacitors are labelled “474”, and are capacitors C11, C43, C44, C45 and C46.



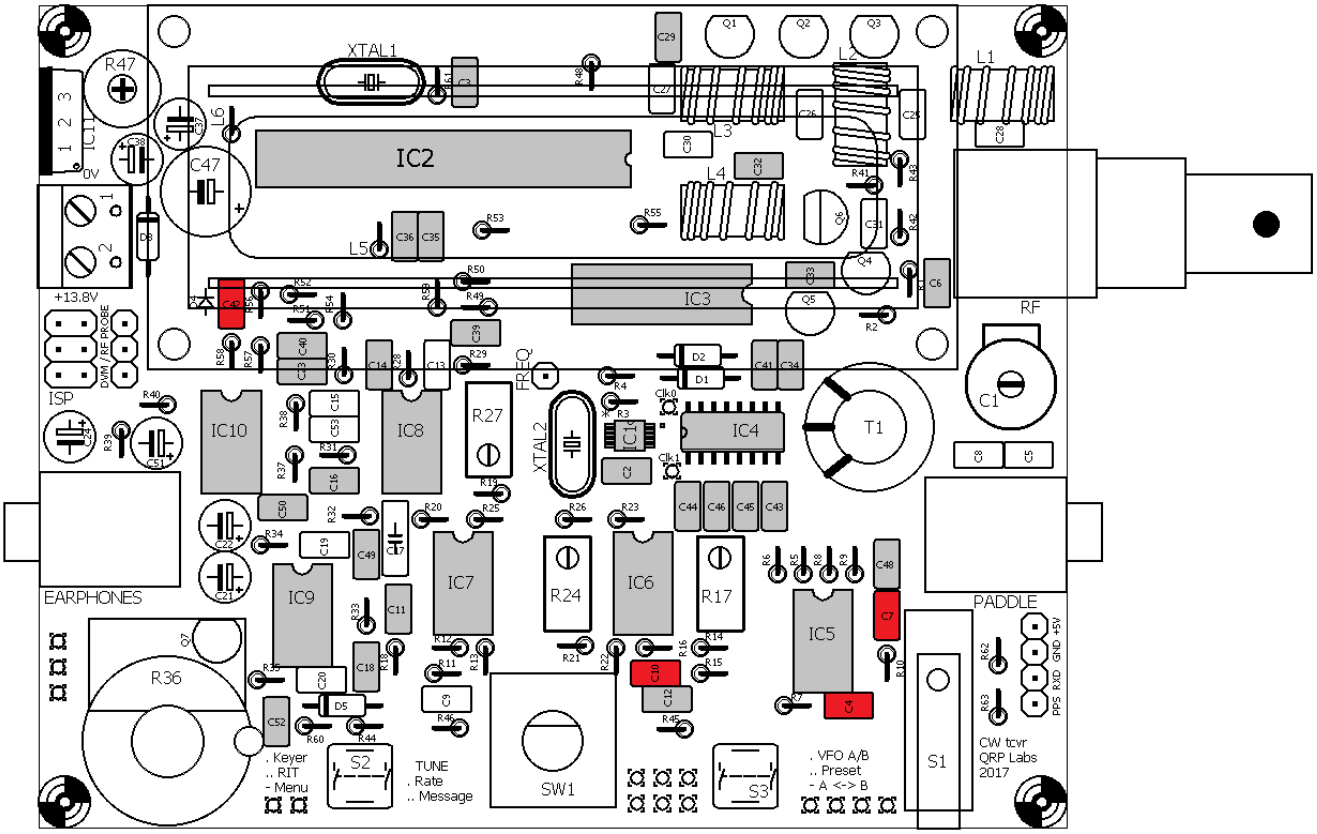
3.6 Install all 1nF, “102” capacitors

The 1nF capacitors are labelled “102” and are capacitors C14, C16, C18, C23 and C33.



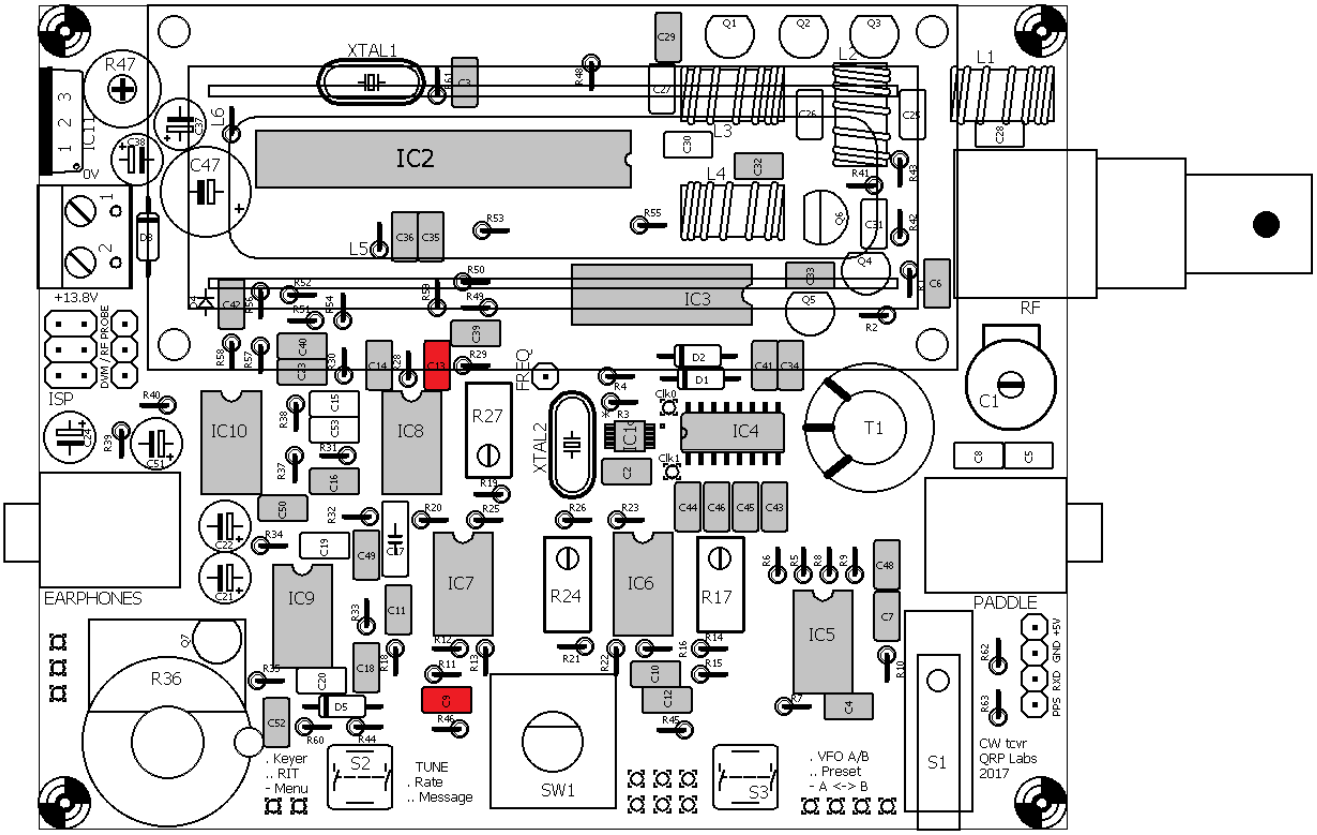
3.7 Install all 10nF, “103” capacitors

The 10nF capacitors are labelled “103” and are capacitors C4, C7, C10 and C42.



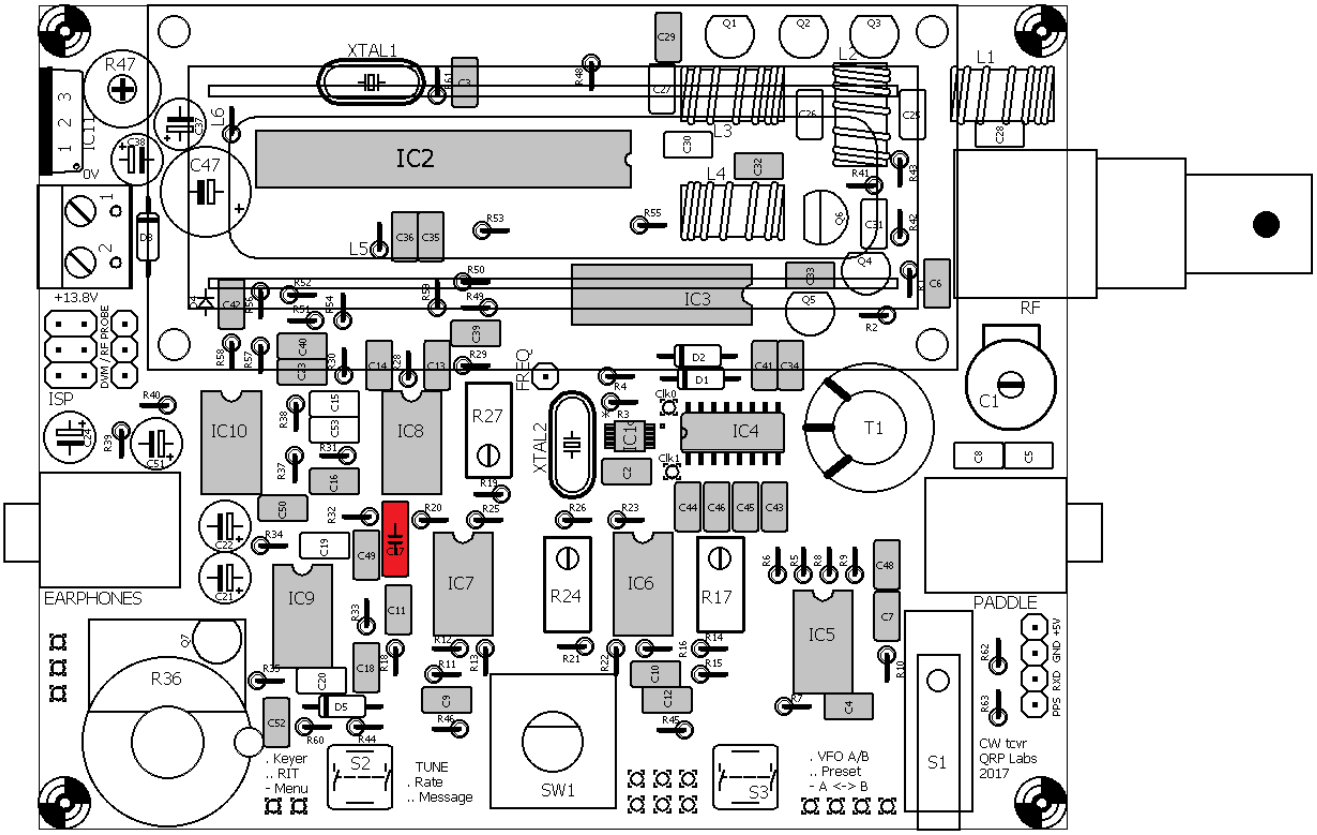
3.8 Install 47nF, “473” capacitors

The 47nF capacitors are labelled “473” and are capacitors C9 and C13.



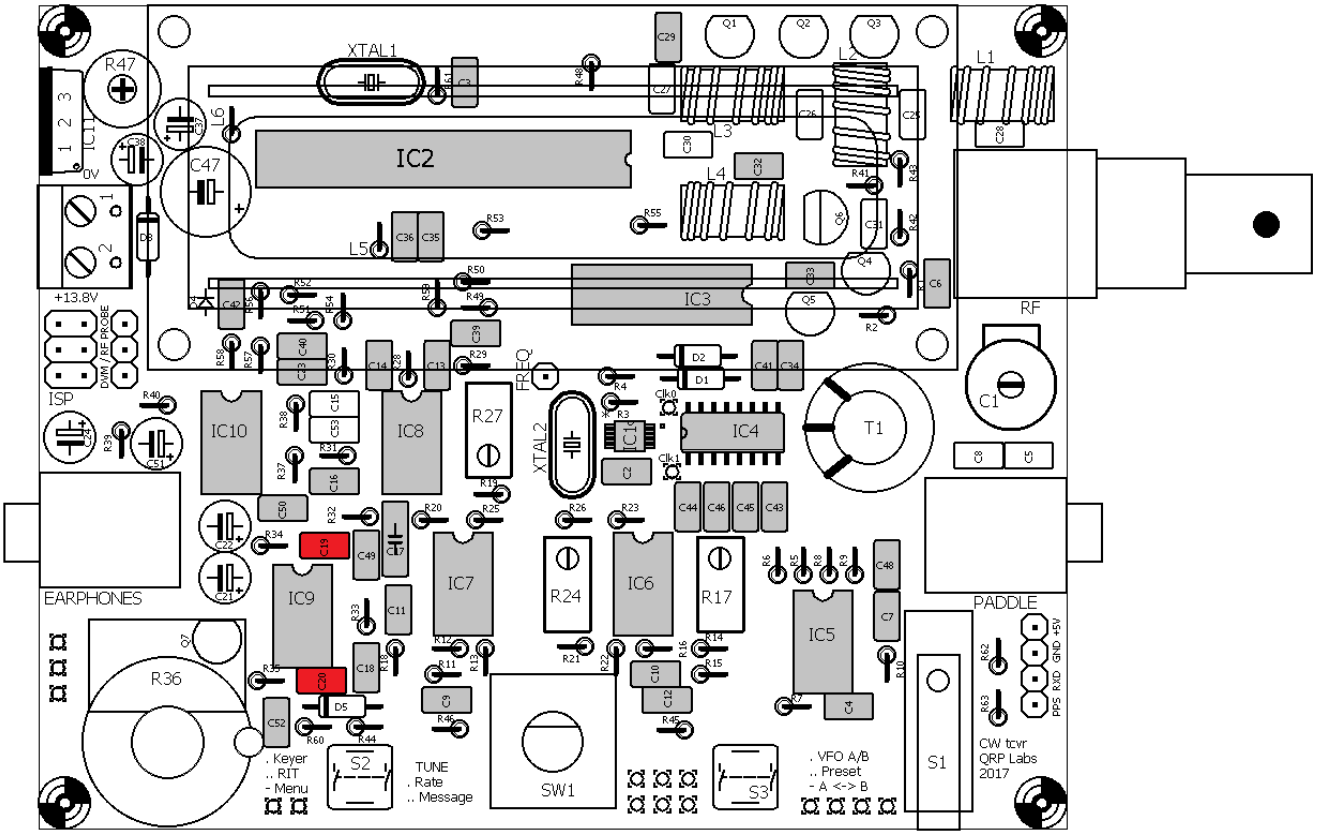
3.9 Install 39nF, “393” capacitor

The 39nF capacitor is labelled “393” and is C17. It has 5mm pin spacing, unlike the others.



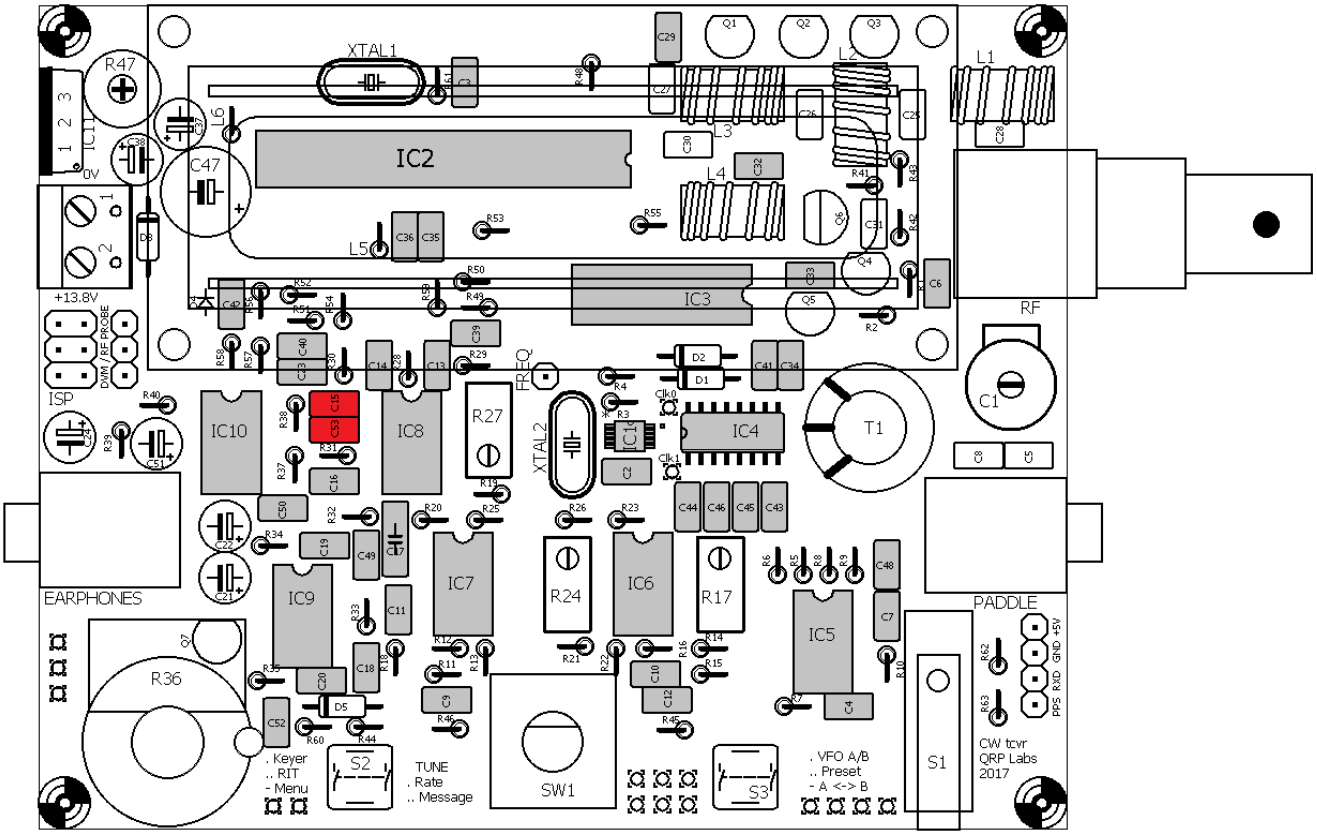
3.10 Install 2.2nF, “222” capacitors

The 2.2nF capacitors are labelled “222” and are C19 and C20.



3.11 Install 33nF, “333” and 3.3nF, “332” capacitors

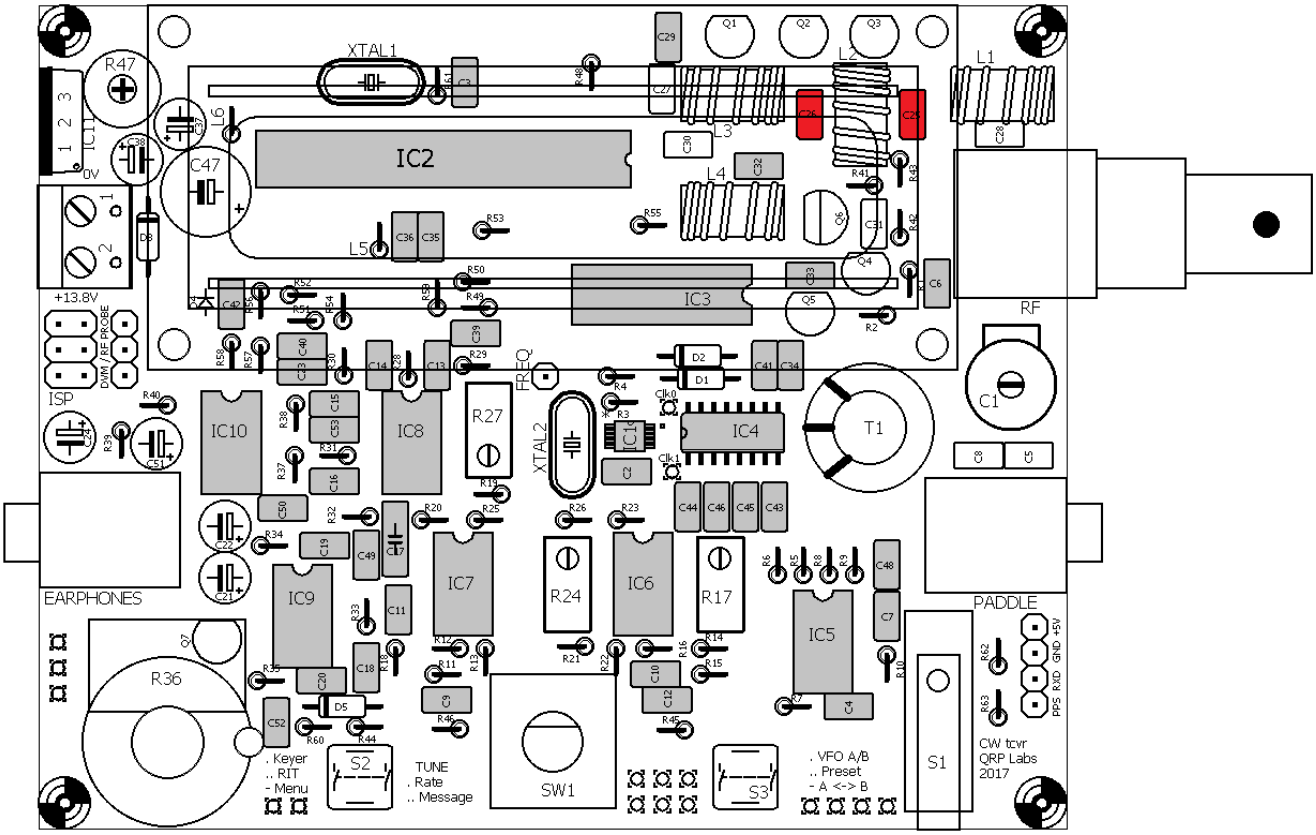
These are labelled “333” and “332”, and in parallel make up 36nF. They are C15 and C53.



3.12 Install capacitors C25 and C26 from the Low Pass Filter kit

The value of these capacitors depends on your chosen band. The capacitors are located inside the separate Low Pass Filter bag in your main kit bag. Refer to the following table to find the correct capacitor value for your band:

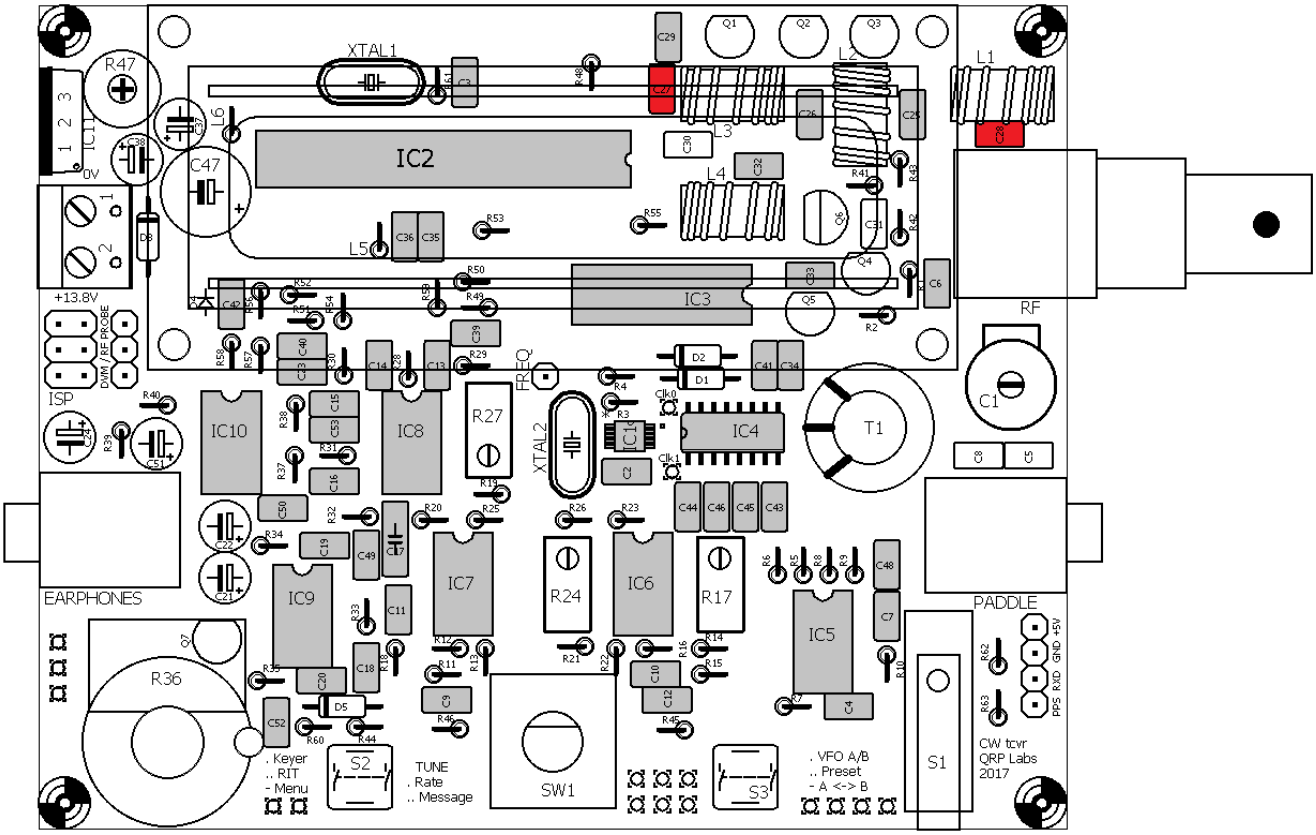
Band	Value	Label
80m	1200pF	“122”
60m	1200pF	“122”
40m	680pF	“681”
30m	560pF	“561”
20m	390pF	“391”
17m	270pF	“271”



3.13 Install capacitors C27 and C28 from the Low Pass Filter kit

The value of these capacitors depends on your chosen bandThe capacitors are located inside the separate Low Pass Filter bag in your main kit bag. Refer to the following table to find the correct capacitor value for your band:

Band	Value	Label
80m	470pF	“471”
60m	680pF	“681”
40m	270pF	“271”
30m	270pF	“271”
20m	180pF	“181”
17m	100pF	“101”



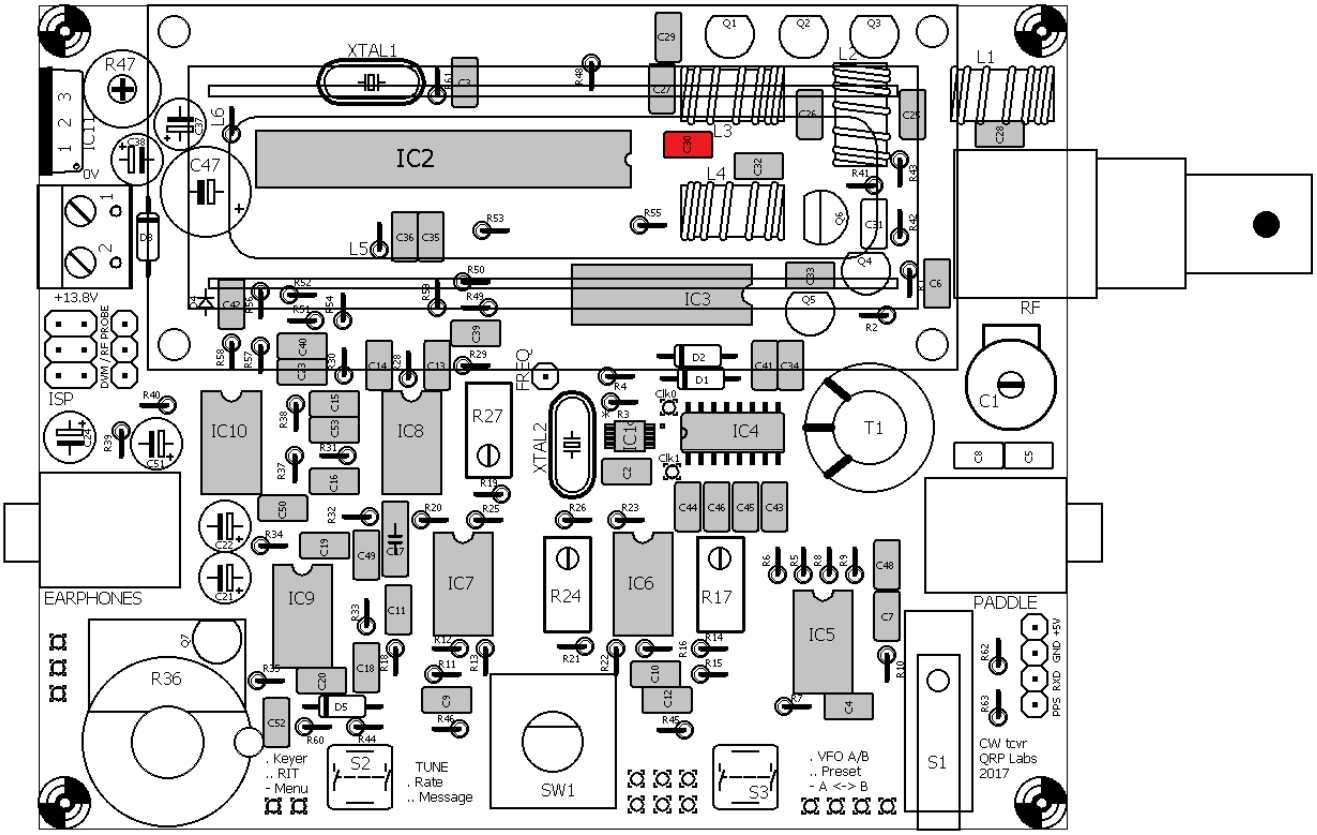
3.14 Install capacitor C30

This capacitor is band dependent. The kit contains all required capacitor values for all bands. Install the one appropriate to your band. Refer to the following table to find the correct capacitor value for your band:

Band	Value	Label
80m	180pF	“181”
60m	30pF 56pF	“300” “560”
40m	56pF	“560”
30m	30pF	“300”
20m	30pF	“300”
17m	30pF	“300”

60m important note: for the 60m band, the capacitor requires both the 30pF and 56pF capacitors to be soldered in parallel, but there is only one component position on the PCB. For 60m, you will need to install one of the capacitors (e.g. 56pF) in the component holes provided, and solder the other one (e.g. 30pF) to the same pads under the PCB. Take care to keep the component wires short and not accidentally touching any other nearby components or soldered pads.

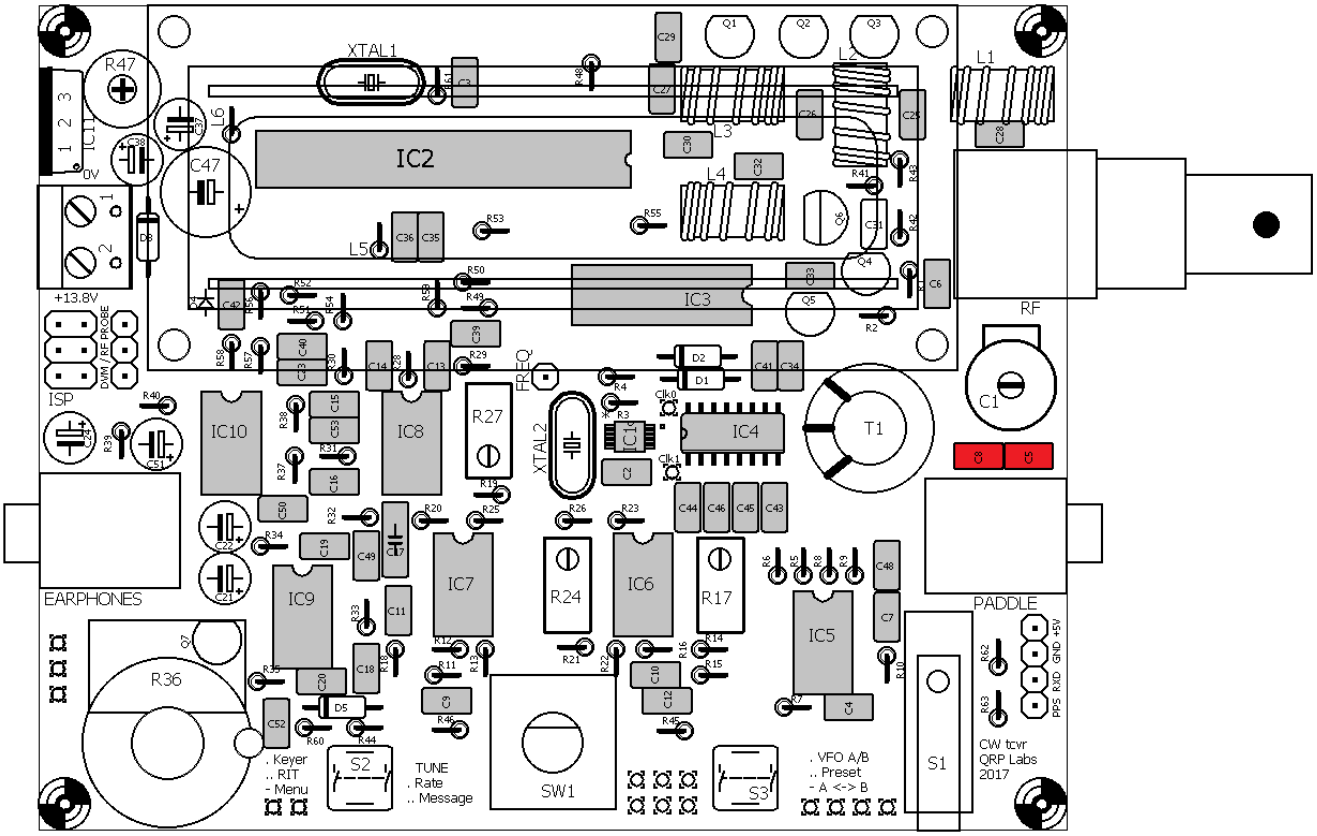
80m important note: the 180pF capacitor supplied has a 0.2-inch (5.08mm) pin spacing. But the holes on the PCB are spaced 0.1-inch (2.5mm). It is necessary to squeeze the capacitors wires carefully closer together to fit the PCB holes.



3.15 Install capacitors C5 and C8

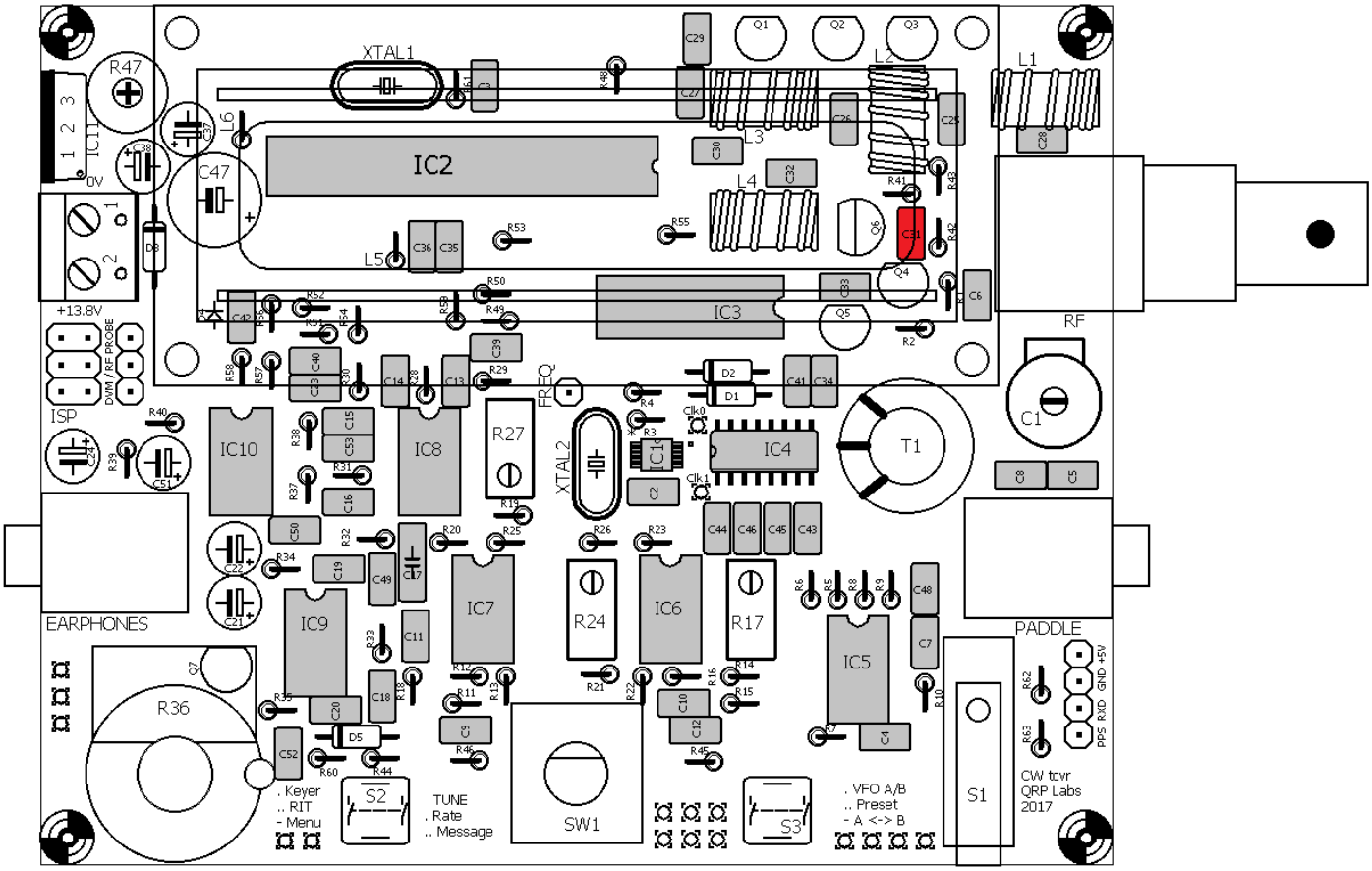
These capacitors are band dependent. They add parallel capacitance to trimmer capacitor C1 to bring it to the required value. The kit contains all required capacitor values. Install the capacitor(s) appropriate to your band. Refer to the following table to find the correct capacitor value(s) for your band. Where “none” is indicated in the table, do not install the corresponding capacitor.

Band	C5 Value	C5 Label	C8 value	C8 Label
80m	39pF	“390”	22pF	“220”
60m	39pF	“390”	22pF	“220”
40m	39pF	“390”	none	
30m	22pF	“220”	none	
20m	none		none	
17m	none		none	



3.16 Install 1uF, “105” capacitor

The 1uF capacitor is labelled “105” and is C31.

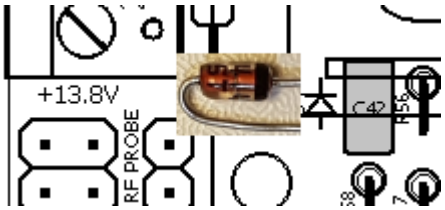


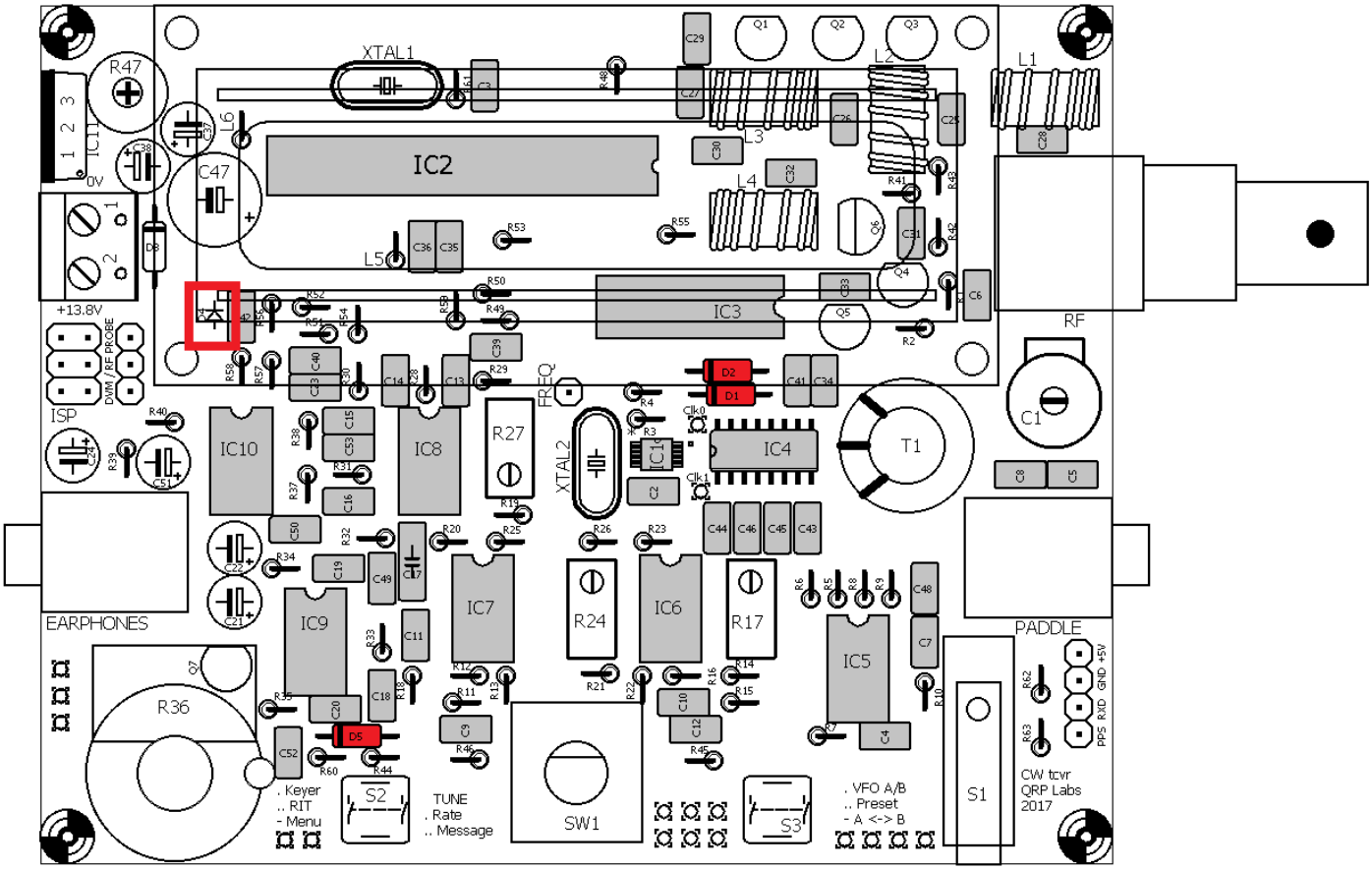
3.17 Install 1N4148 diodes

There are 4 small glass red-ish coloured diodes, D1, D2, D4 and D5. Note that D1, D2 and D5 are installed flat on the PCB (see photo, right). D4 is installed vertically, (see photo, far right).

In all cases it is critical to install the diodes the right way around, these are a polarised component and must only be installed in the PCB with the correct orientation! In the case of D1, D2 and D5, the PCB silkscreen printing shows a white stripe at the end of the diode body which must match the black stripe on the left end of the diode (photo above).

For D4, the black stripe on the diode body must be orientated with the horizontal bar of the diode symbol on the layout diagram (see photo, right). D4 is mounted vertically. The location of D4 is indicated by the red square, near the top left of the diagram on the following page.



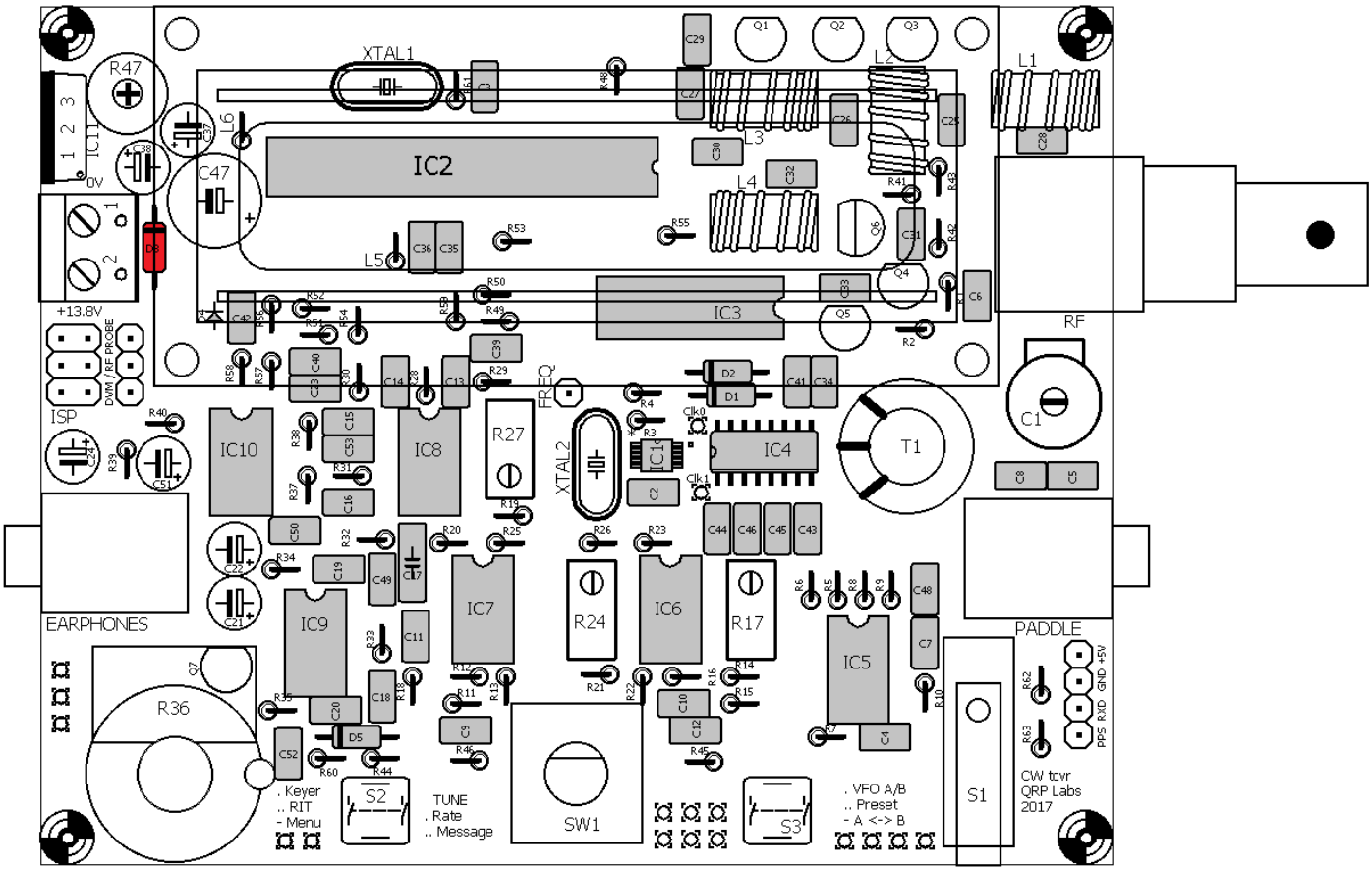


3.18 Install 1N5819 diode

This diode D3 is the larger diode with a black body, and a white stripe. It is installed horizontally. Again, it must be orientated correctly, with the white stripe on the diode matching the white stripe on the PCB.

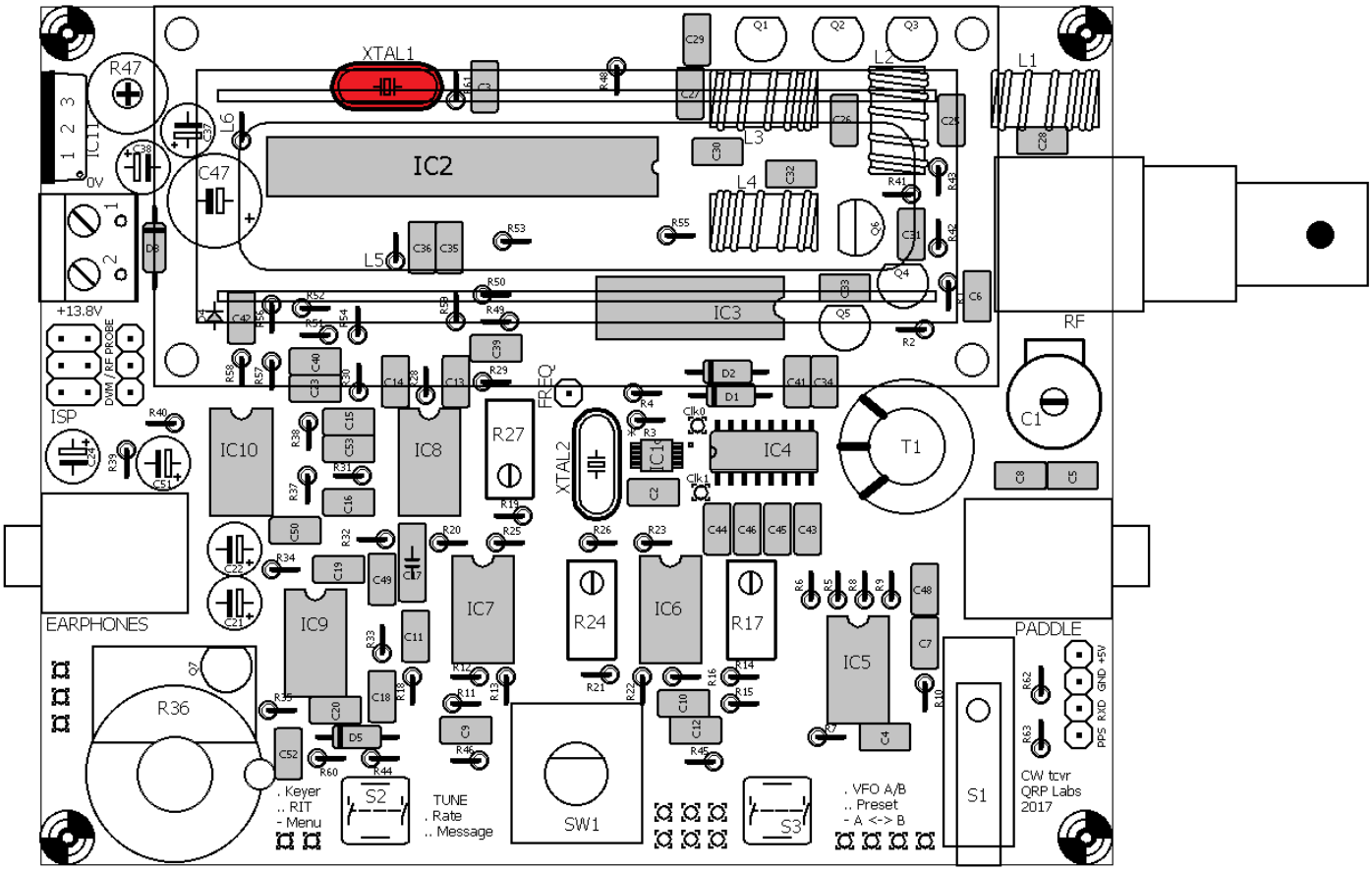
This diode protects the radio against reverse polarity. So that if you connect the power to the board the wrong way around, you do not destroy it. A Schottky diode is used because the forward conducting voltage drop is less than an ordinary diode. However, on transmit, the voltage drop across this diode can still be up to 400mV. This reduction in voltage does slightly decrease the output power.

IF you want to squeeze every last milliwatt of output power out of the radio, and IF you trust yourself NEVER to connect the power supply in reverse by mistake, then if you just install a jumper wire instead of D3, it would give you a little higher RF power output.



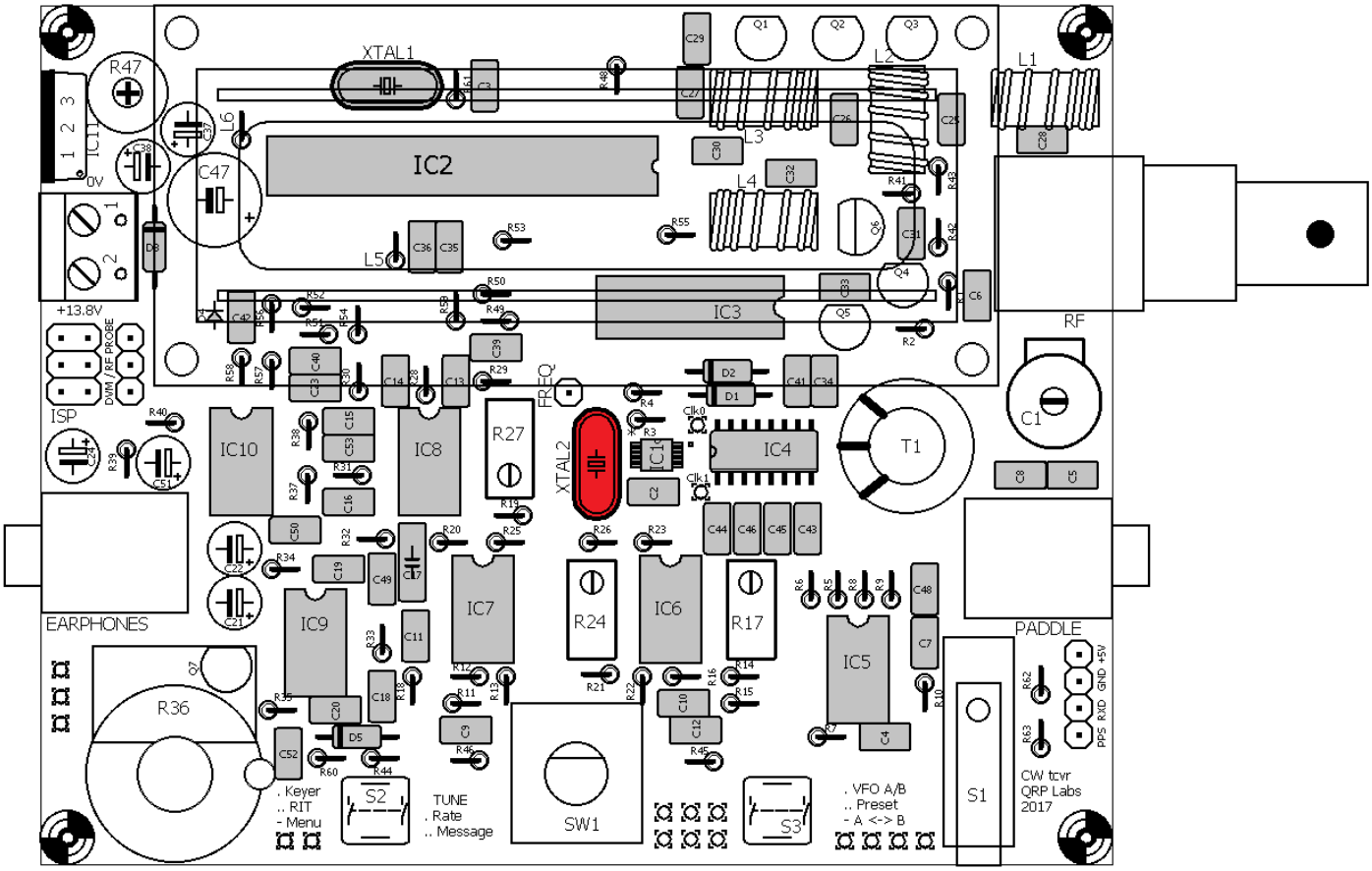
3.19 Install 20MHz crystal XTAL1

The engraving on this crystal is “20.000”.



3.20 Install 27MHz crystal XTAL2

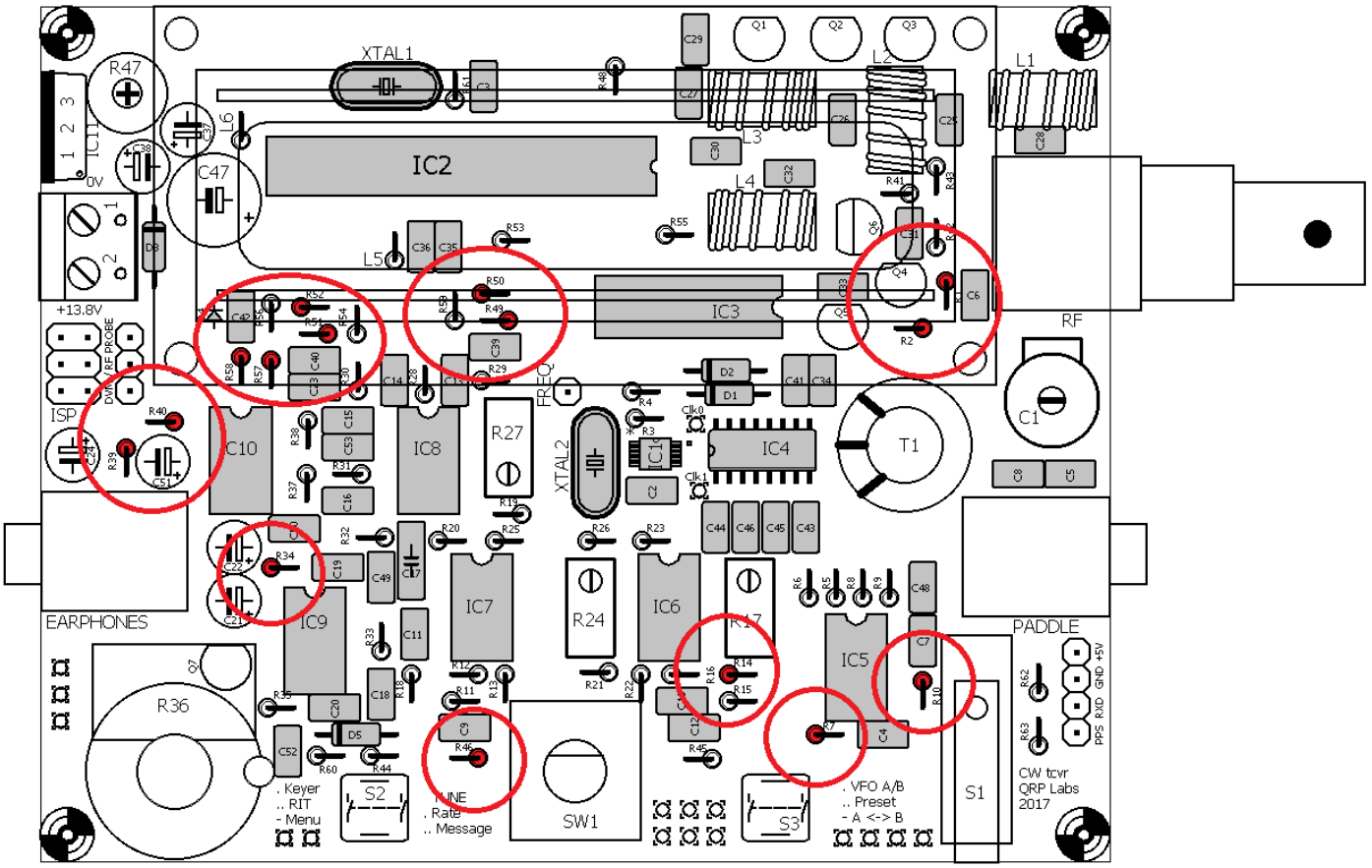
The engraving on this crystal is “27.000”.



3.21 Install all 10K resistors

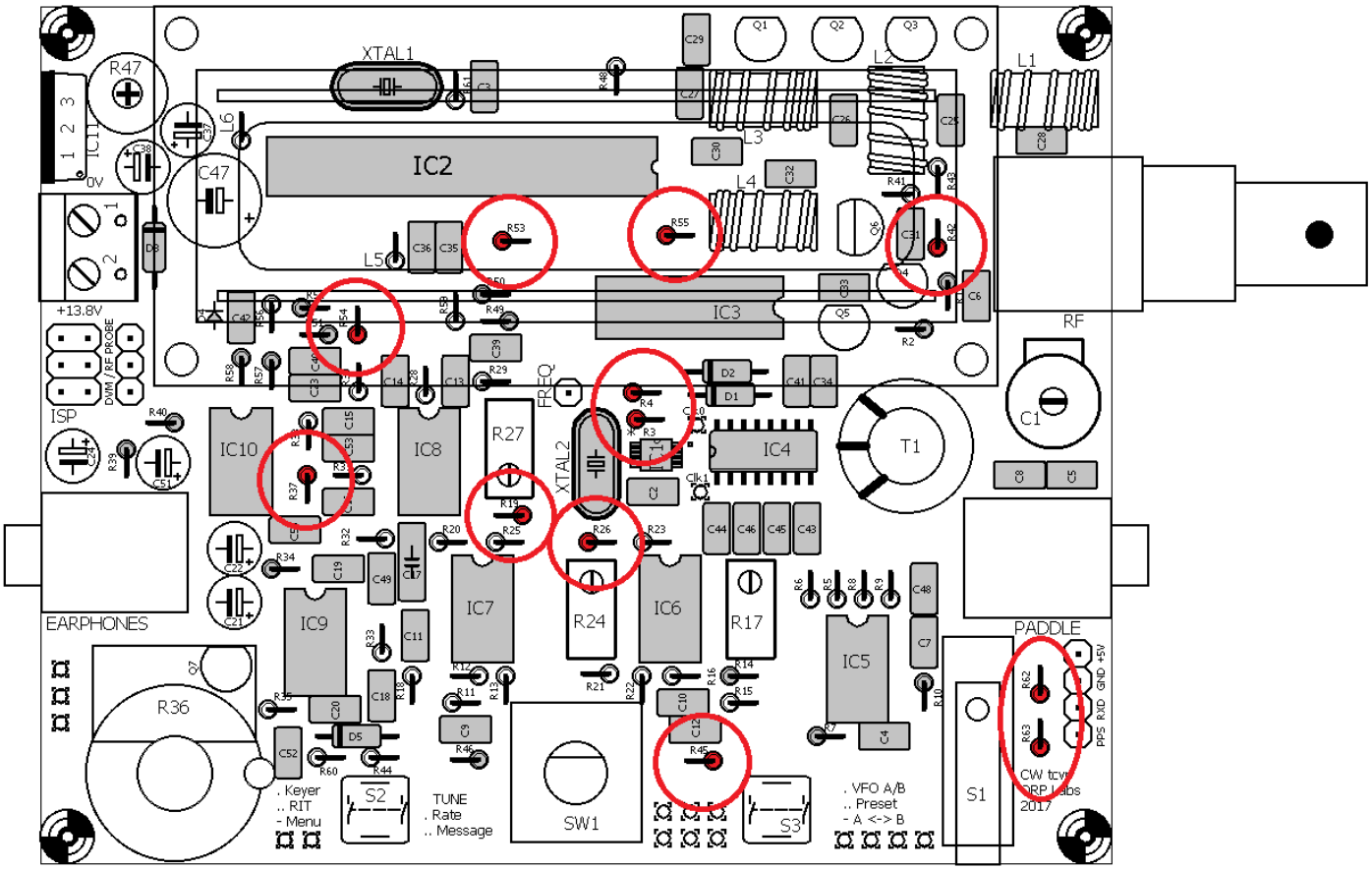
There are 15 10K resistors in the kit, these are R1, R2, R7, R10, R14, R34, R39, R40, R46, R49, R50, R51, R52, R57 and R58. The colour code is brown-black-black-red-brown. All of the resistors in the kit are installed vertically (see photo, right).

Resistors are not polarised components and it does not matter which way around they are installed. For simplicity I recommend installing the body end of the resistor next to the circle on the PCB silkscreen (and the layout diagrams).



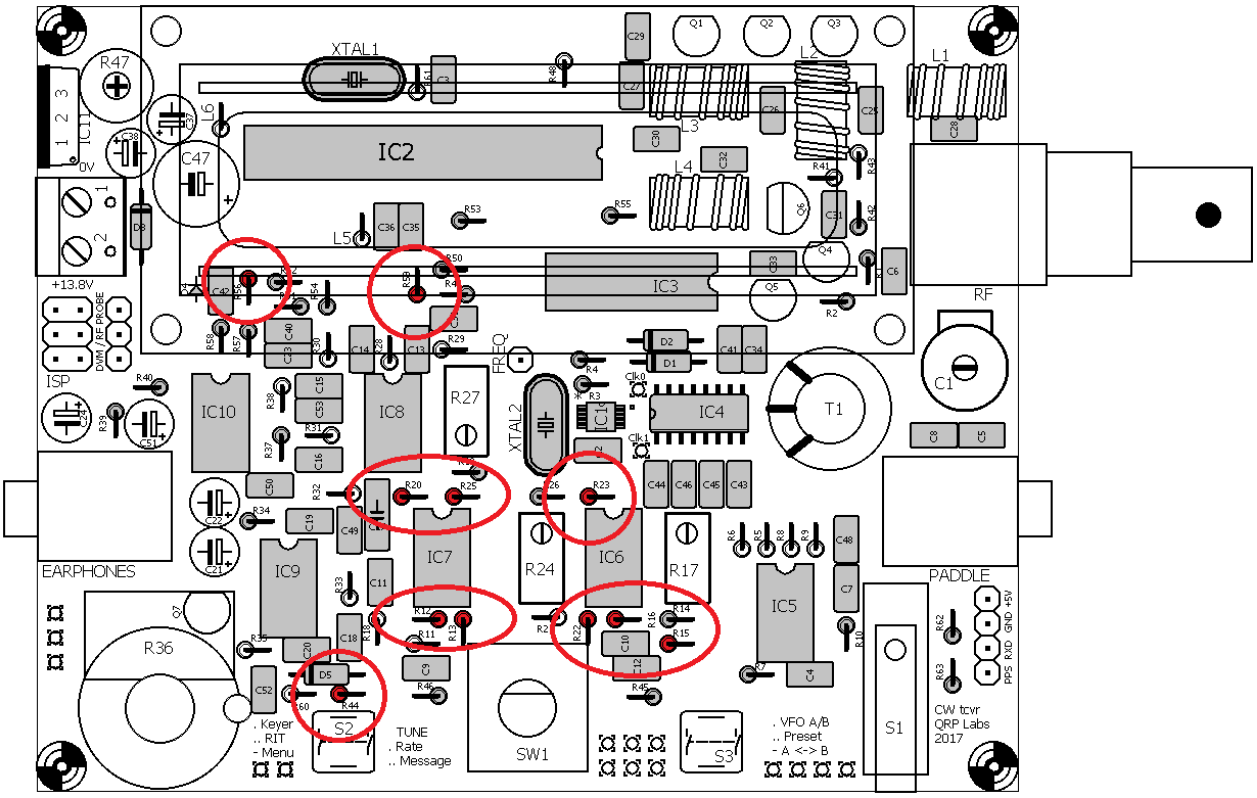
3.22 Install all 1K resistors

There are 12 1K resistors in the kit, these are R3, R4, R19, R26, R37, R42, R45, R53, R54, R55, R62 and R63. The colour code is brown-black-black-brown-brown.



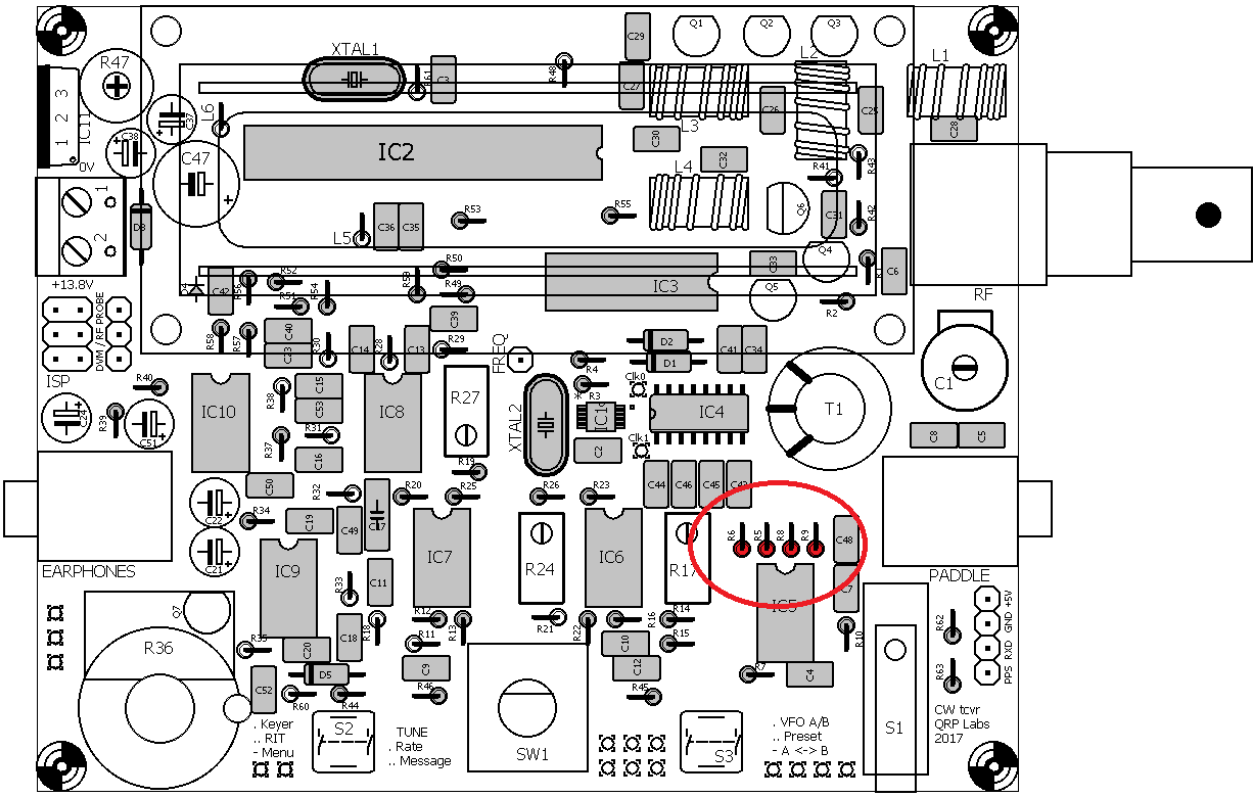
3.23 Install all 3.3K resistors

There are 11 3.3K resistors in the kit, these are R12, R13, R15, R16, R20, R22, R23, R25, R44, R56 and R59. The colour code is orange-orange-black-brown-brown.



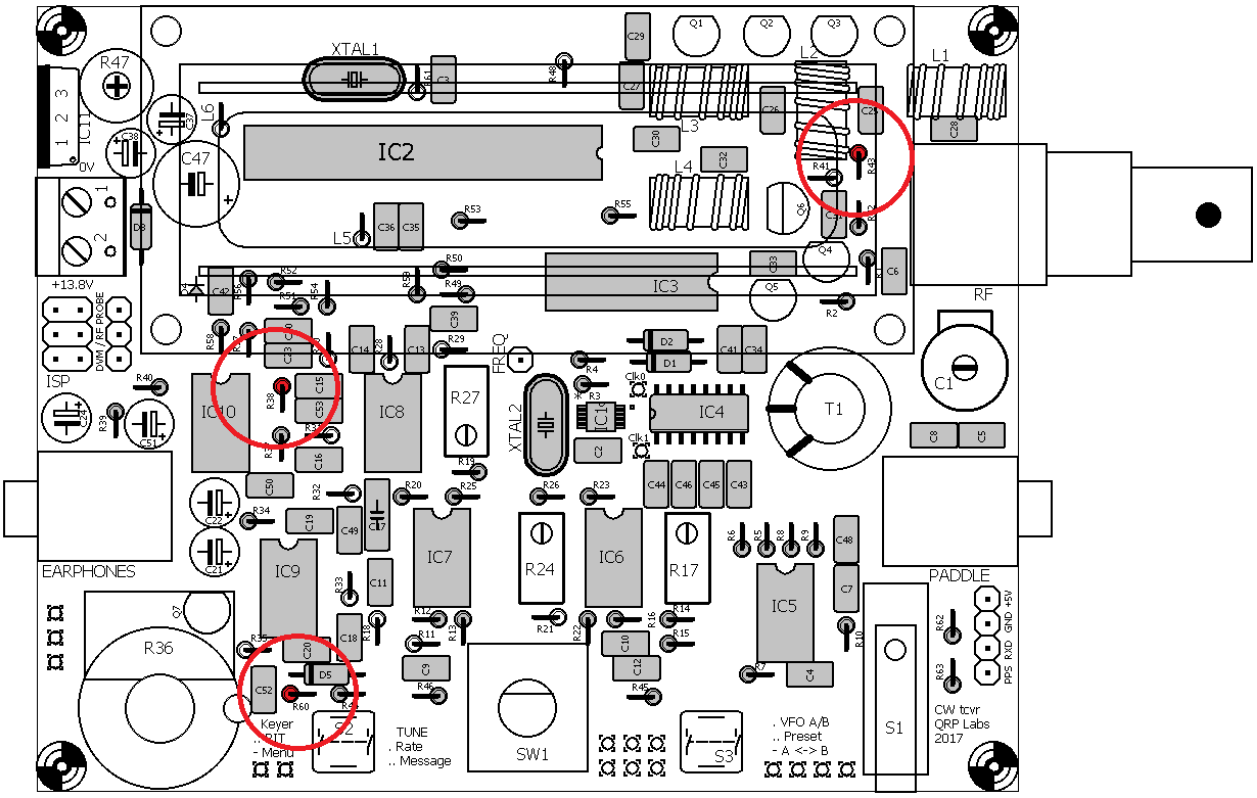
3.24 Install all 100-ohm resistors

There are four 100-ohm resistors in the kit, these are R5, R6, R8 and R9. The colour code is brown-black-black-black-brown.



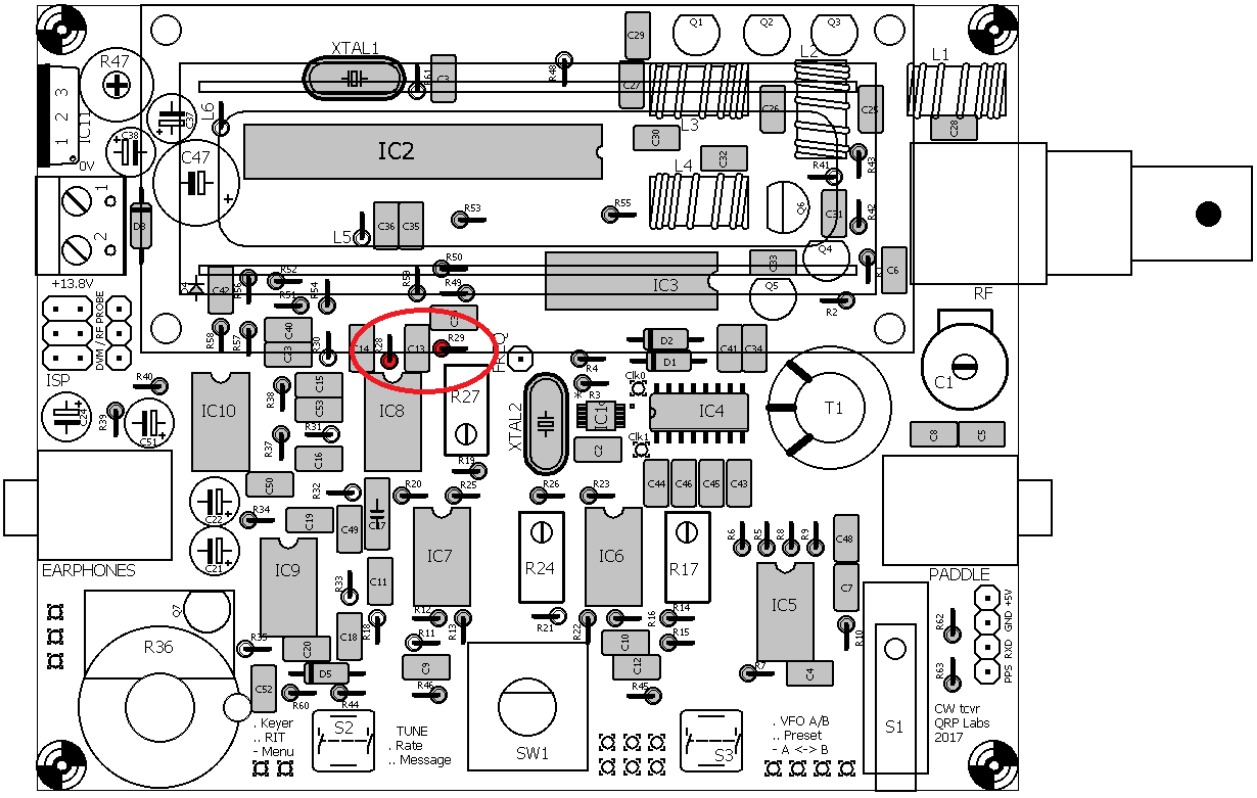
3.25 Install 120K resistors

There are three 120K resistors in the kit, these are R38, R43 and R60. The colour code is brown-red-black-orange-brown.



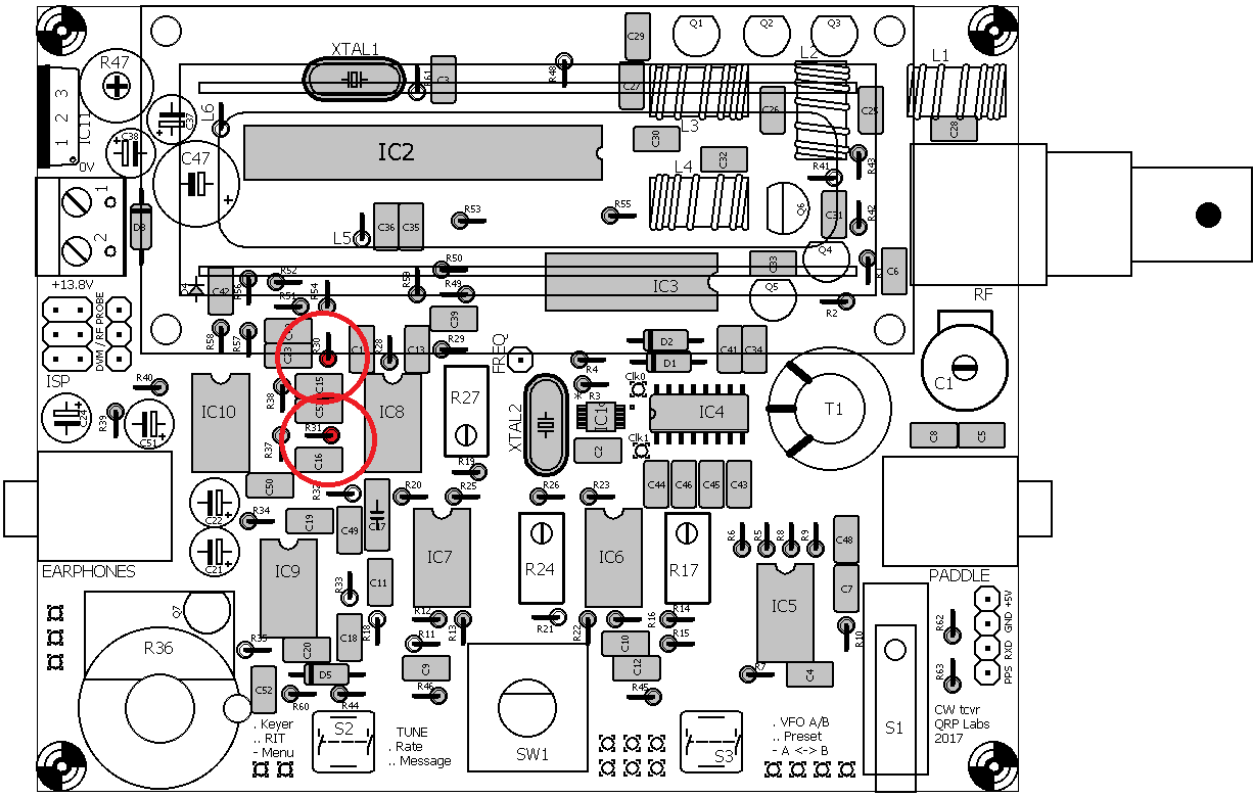
3.26 Install 33K resistors

There are two 33K resistors in the kit, these are R28 and R29. The colour code is orange-orange-black-red-brown.



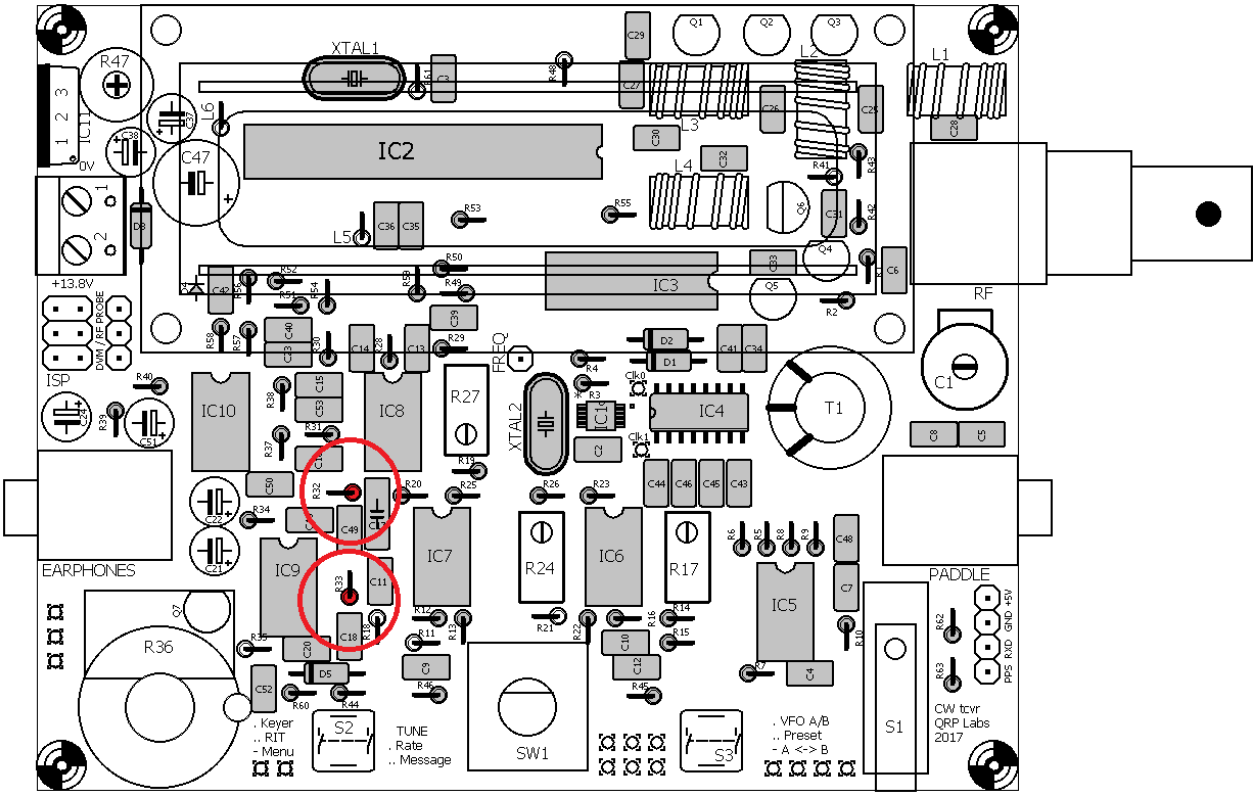
3.27 Install 47K resistors

There are two 47K resistors in the kit, these are R30 and R31. The colour code is yellow-purple-black-red-brown.



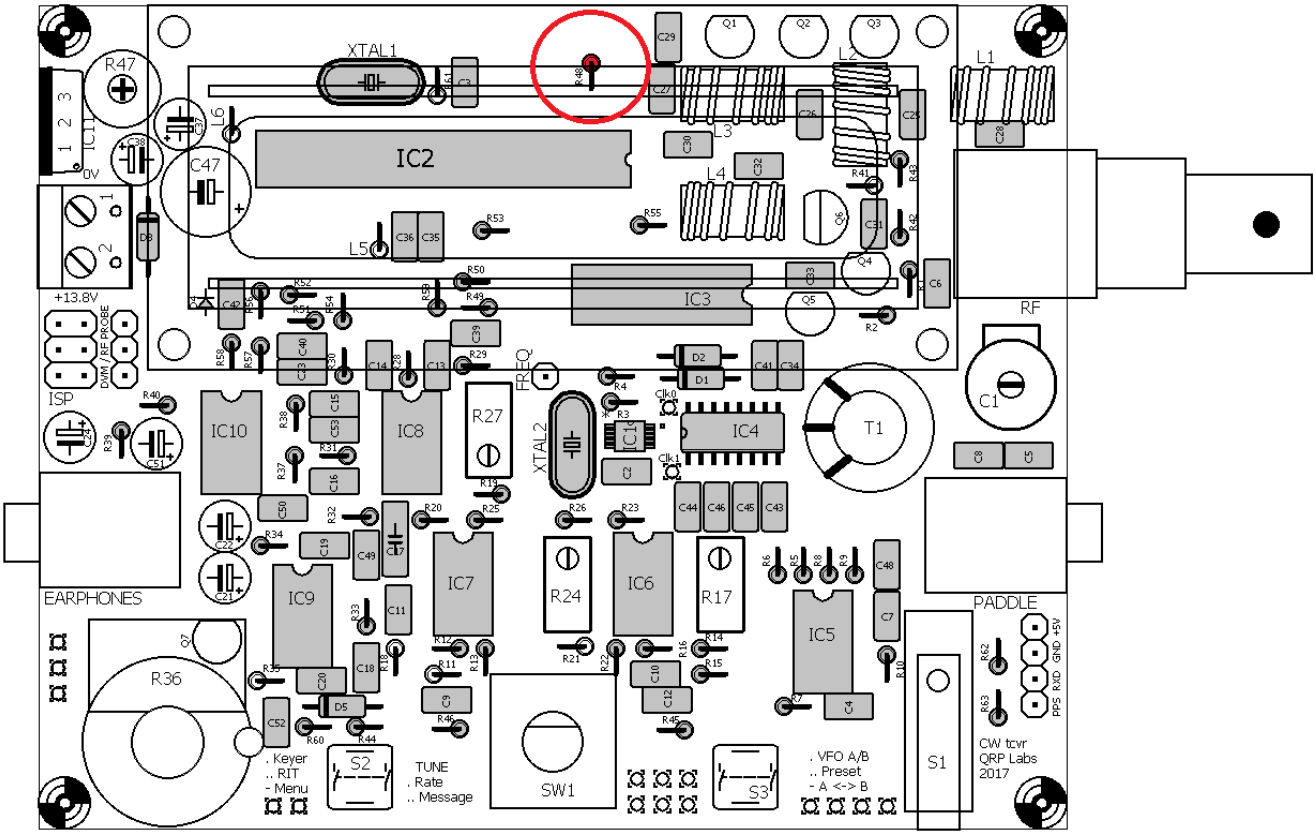
3.28 Install 36K resistors

There are two 36K resistors in the kit, these are R32 and R33. The colour code is orange-blue-black-red-brown.



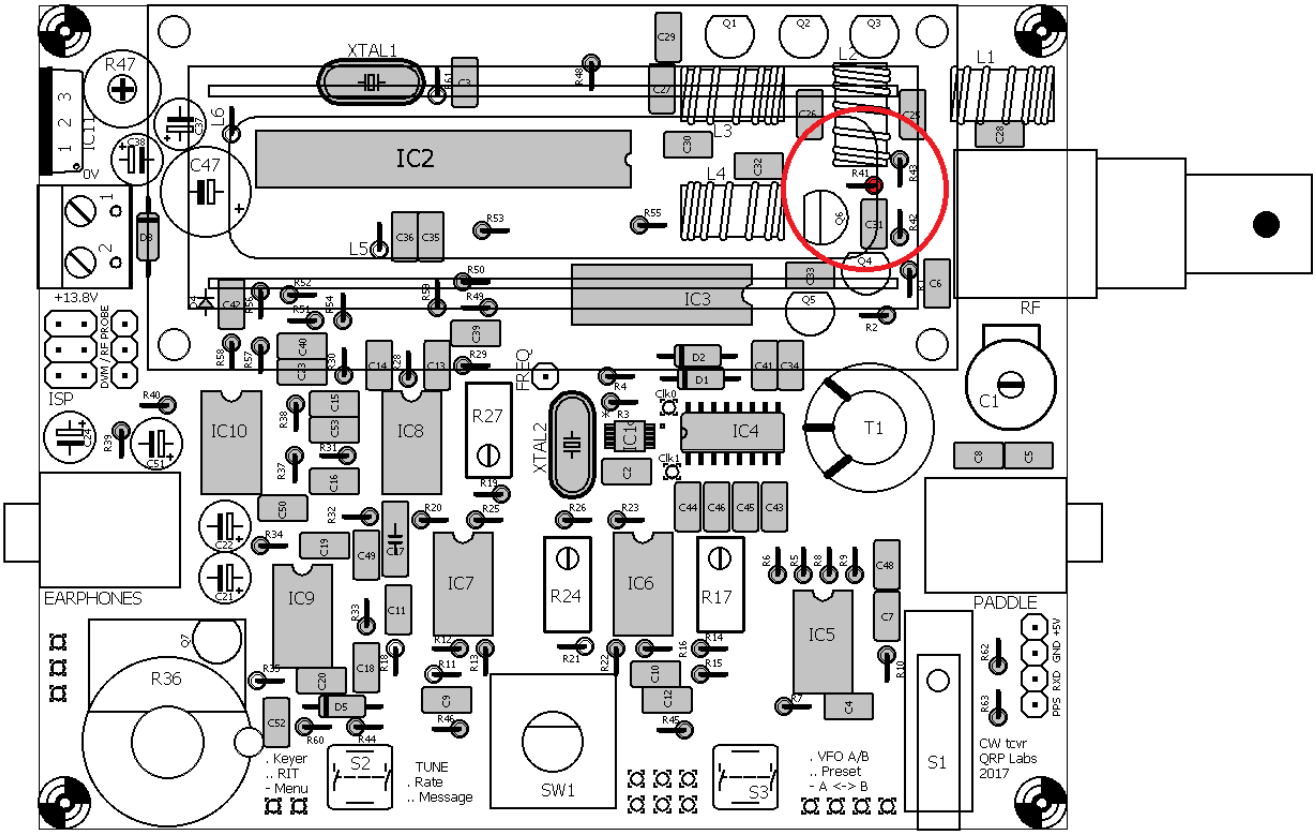
3.29 Install 270-ohm resistor R48

This resistor has colour code red-purple-black-black-brown.



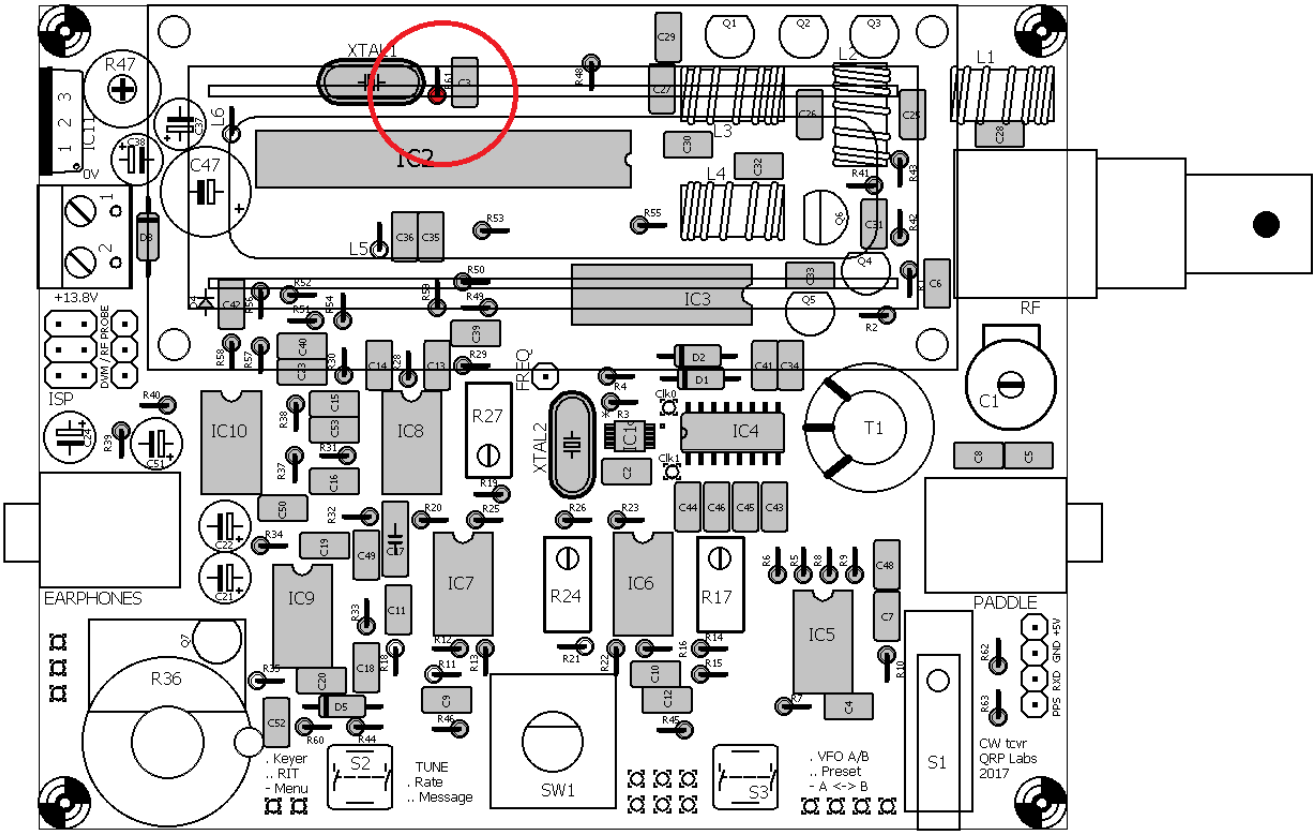
3.30 Install 470-ohm resistor R41

This resistor has colour code yellow-purple-black-black-brown.



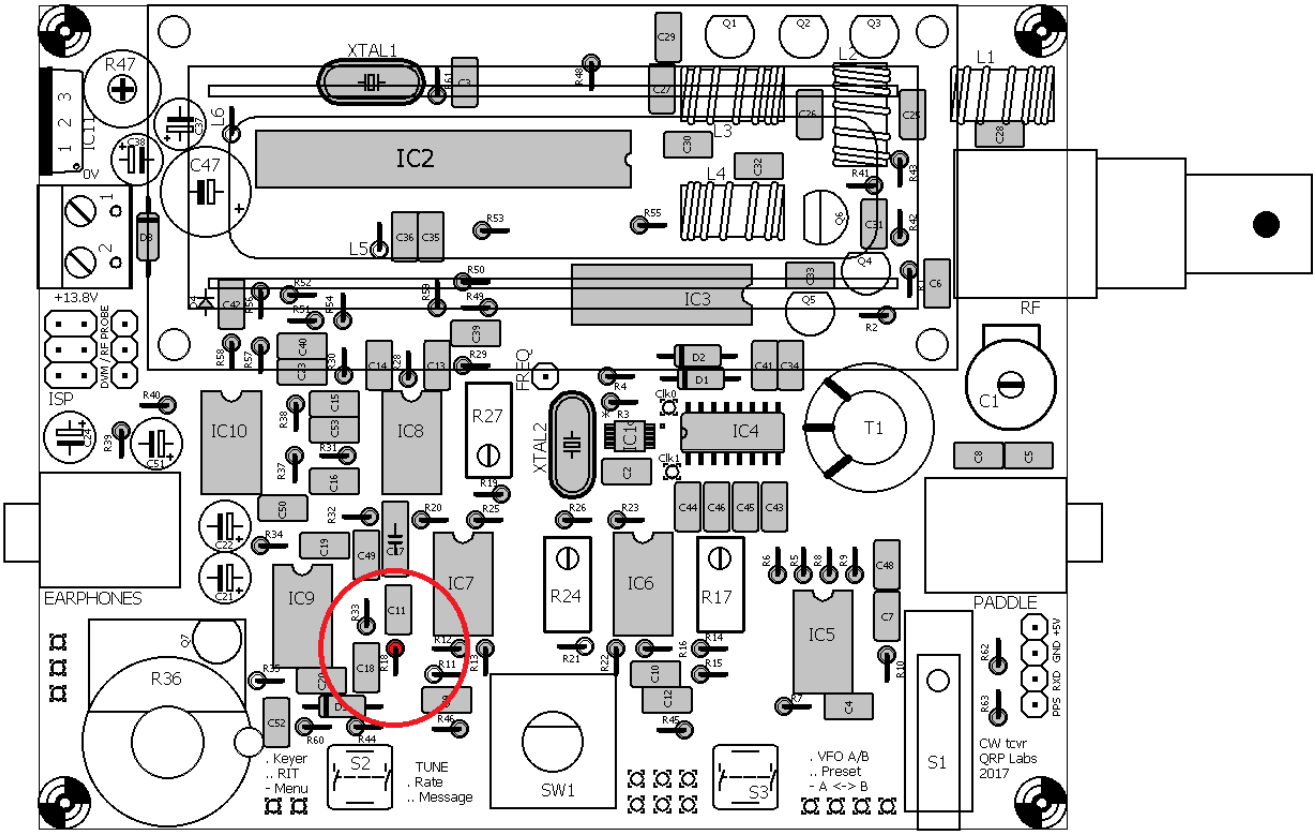
3.31 Install 3.9K resistor R61

This resistor has colour code orange-white-black-brown-brown.



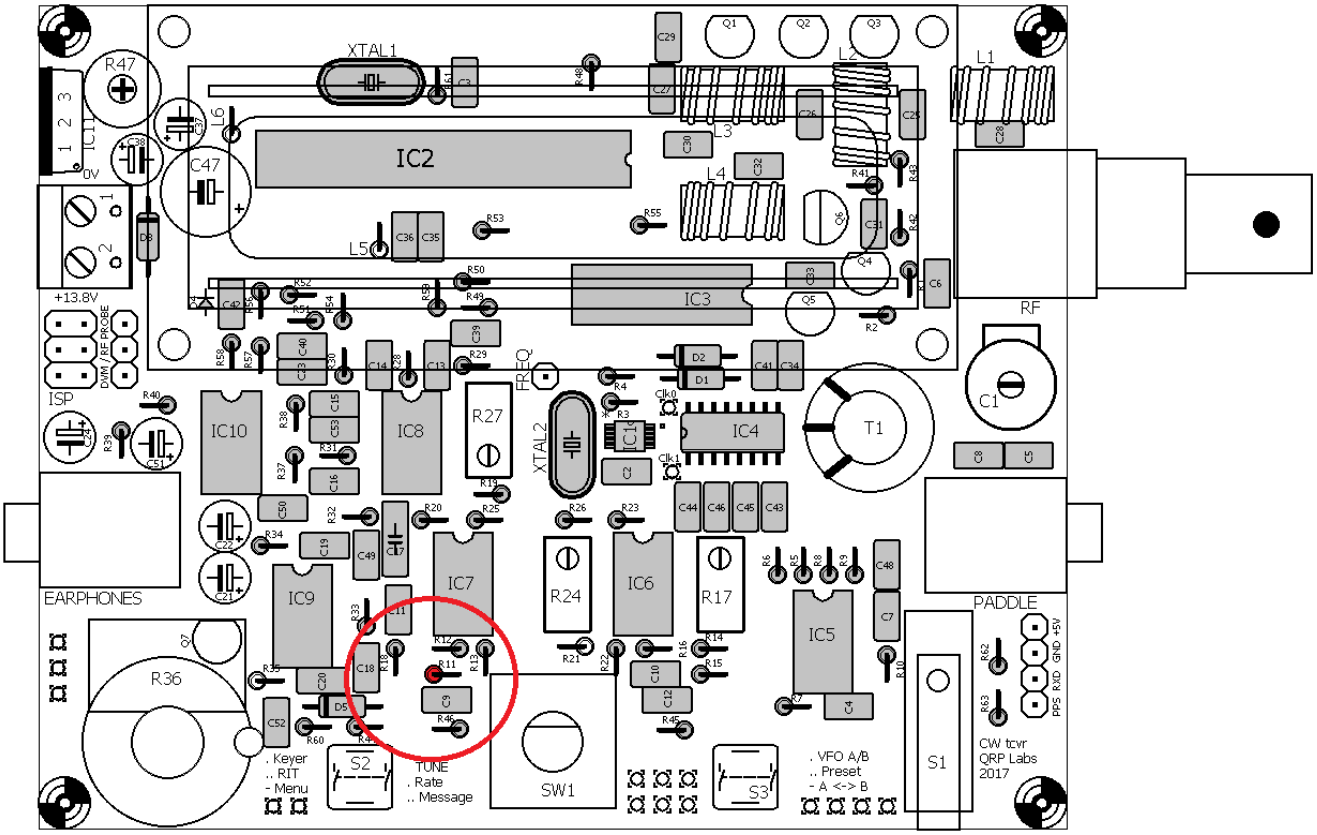
3.32 Install 4.3K resistor R18

This resistor has colour code yellow-orange-black-brown-brown.



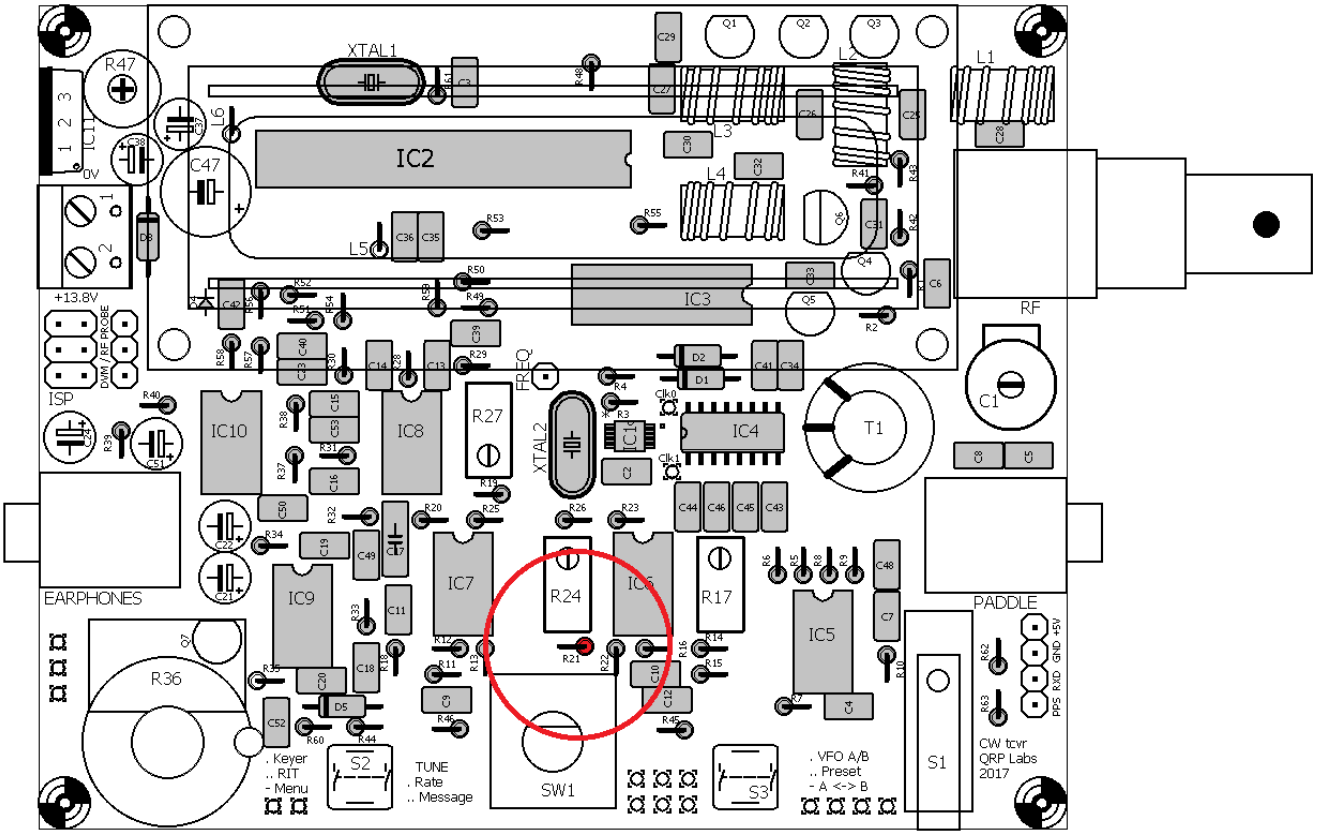
3.33 Install 5.1K resistor R11

This resistor has colour code green-brown-black-brown.



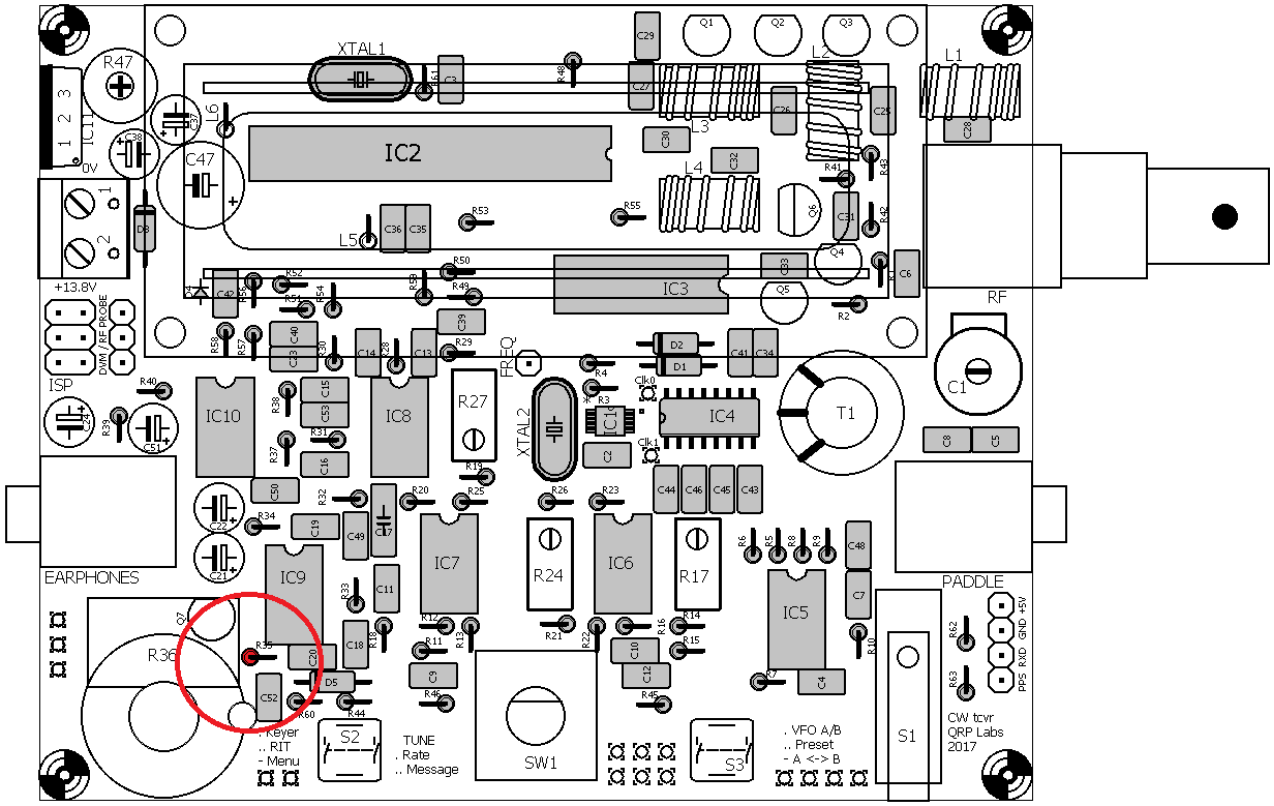
3.34 Install 7.5K resistor R21

This resistor has colour code purple-green-black-brown-brown.



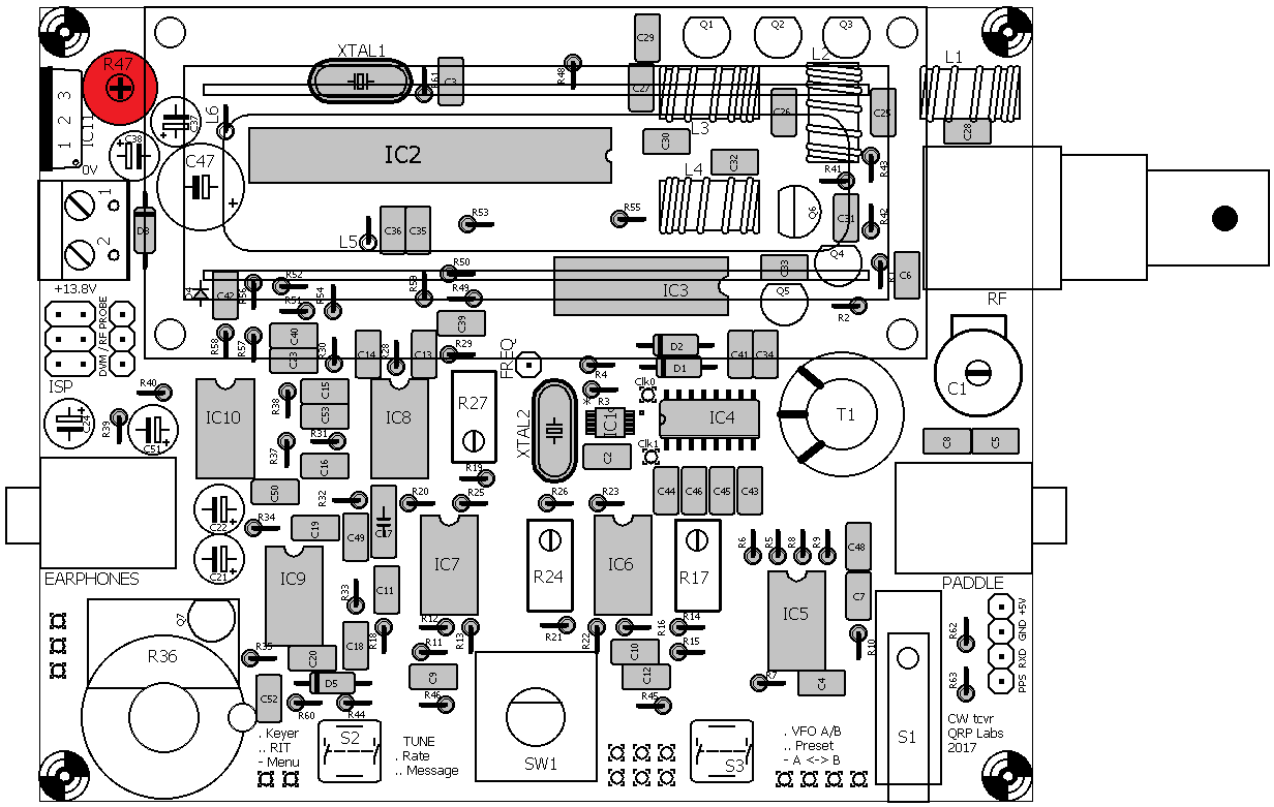
3.35 Install 750K resistor R35

This resistor has colour code purple-green-black-orange-brown.



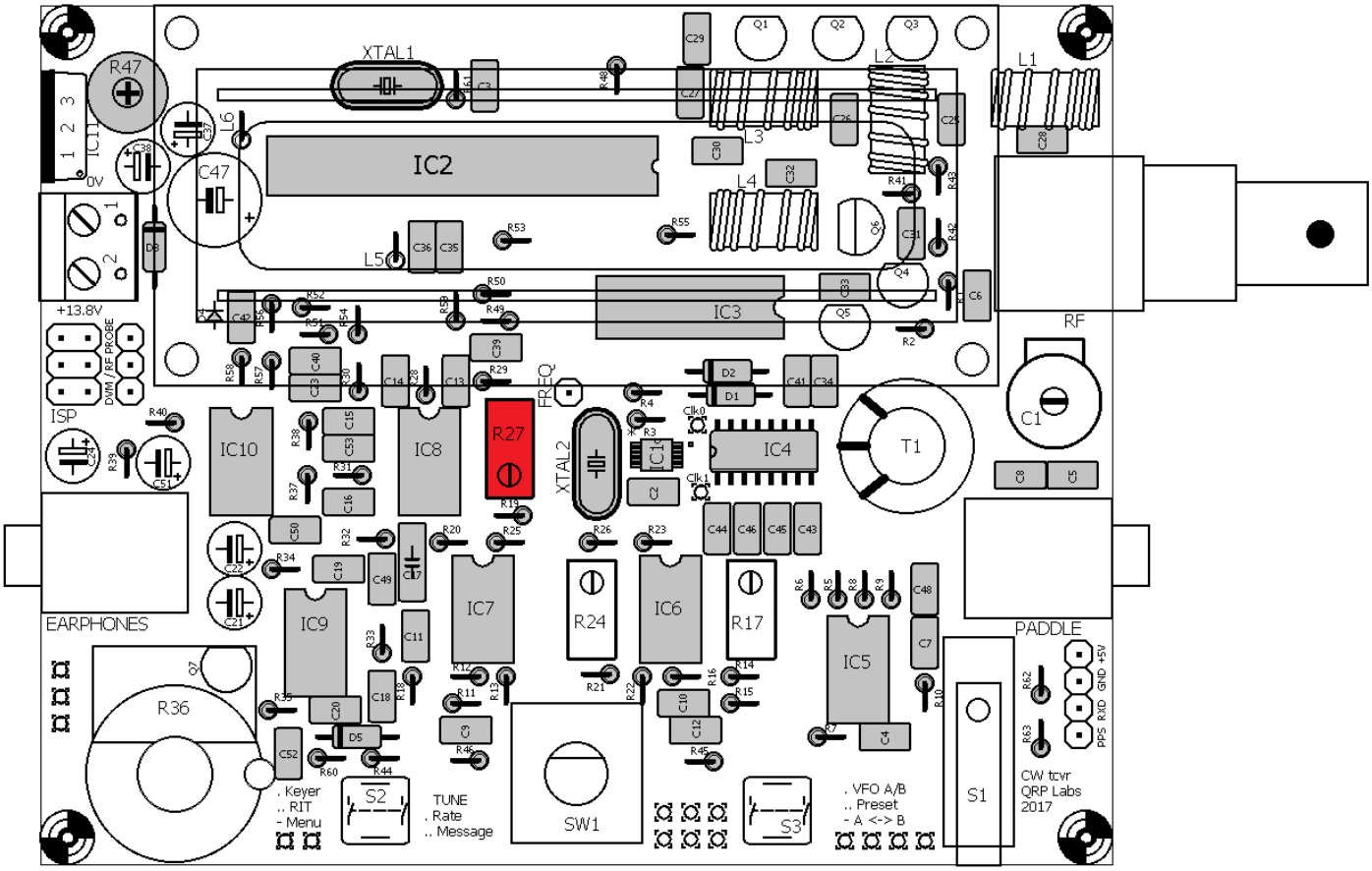
3.36 Install 100K trimmer potentiometer R47

This is the LCD contrast potentiometer. The label on R47 is “104”. Carefully apply firm pressure to install the component pins in the holes without bending them.



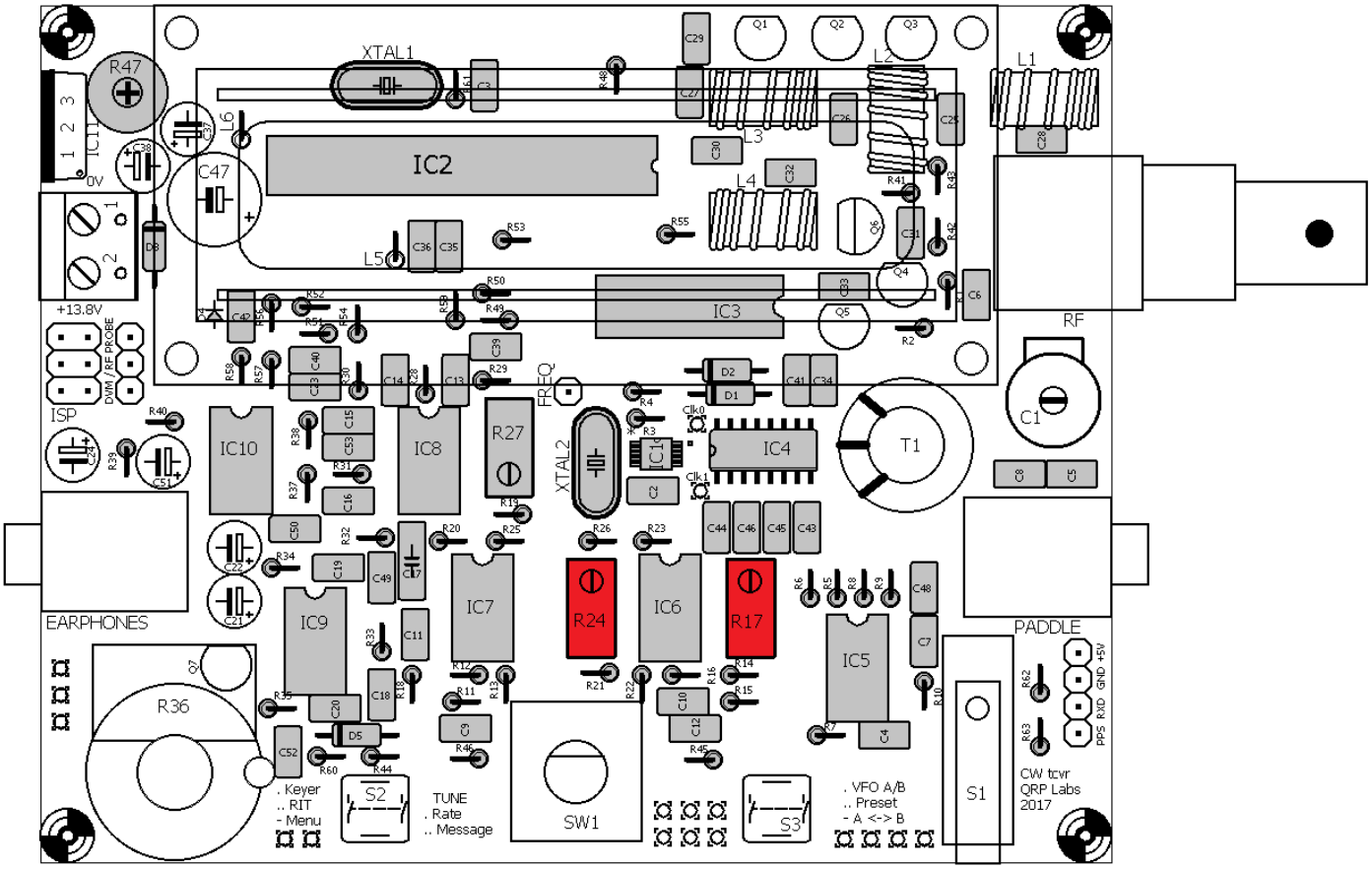
3.37 Install 500-ohm multi-turn trimmer potentiometer

This resistor is the small blue box component with label “501”. It is R27. Make sure that the orientation is correct. **The screw on the resistor must match the screw on the PCB silkscreen and layout diagram.**



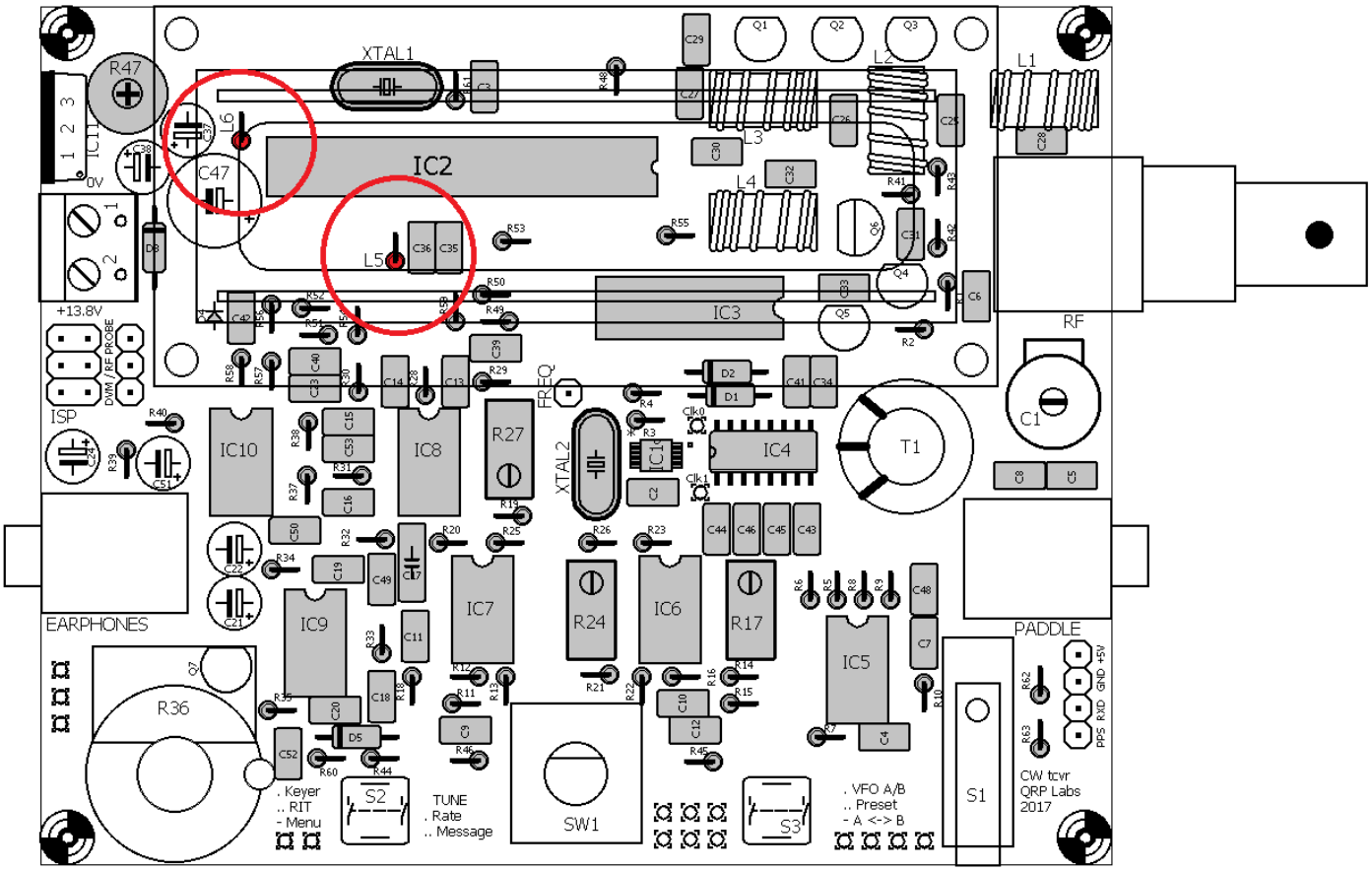
3.38 Install 50K multi-turn trimmer potentiometers

There are two 50K multi-turn trimmer potentiometers, R17 and R24. They are the small blue box components with label “’503”. Make sure that the orientations are correct. **The screws on the resistors must match the screws on the PCB silkscreen and layout diagram.**



3.39 Install two 100uH inductors

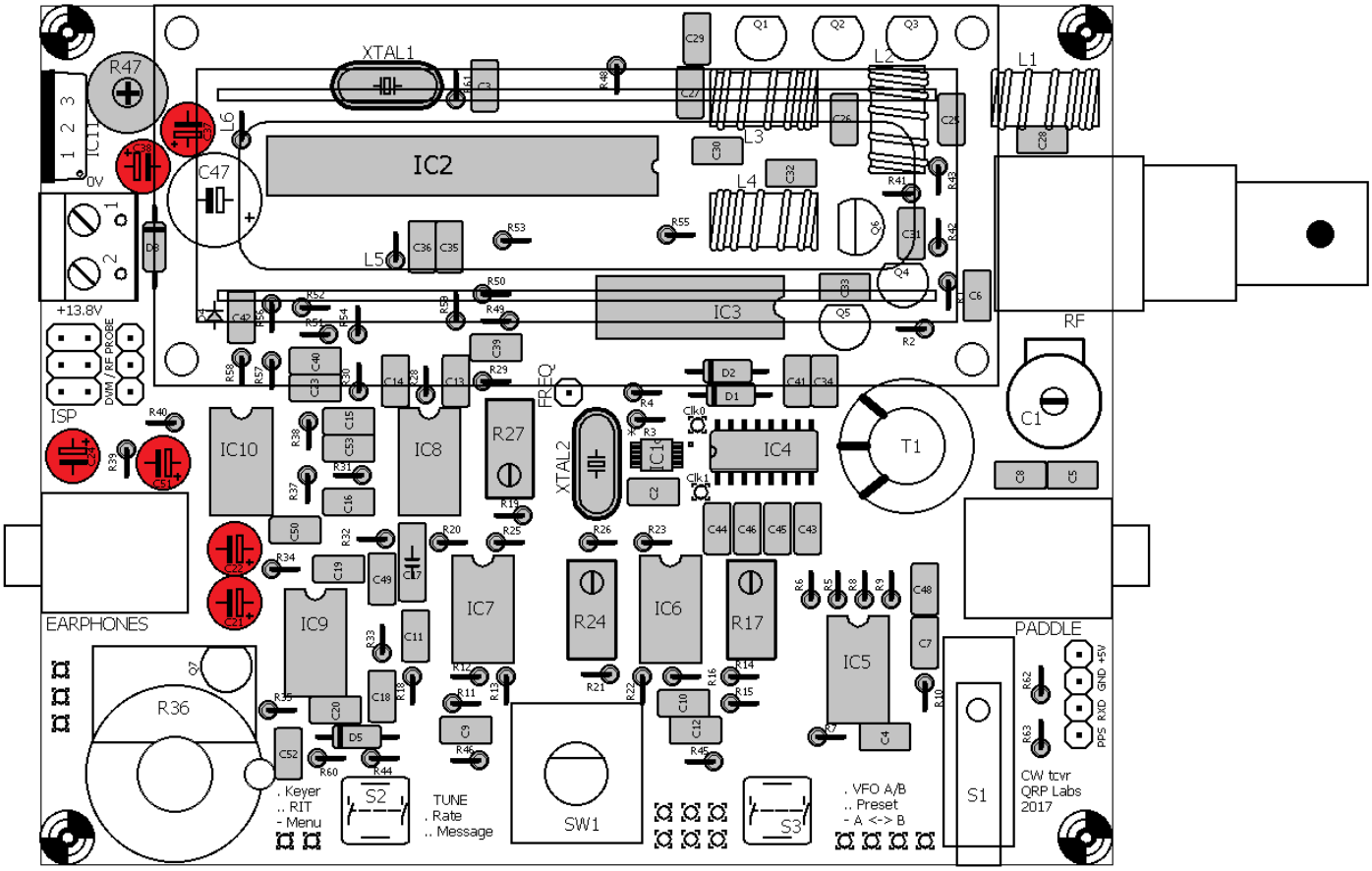
There are two 100uH inductors, L5 and L6. These look like short fat resistors. They are installed vertically on the PCB, in the same way as the resistors are installed.



3.40 Install 10uF capacitors

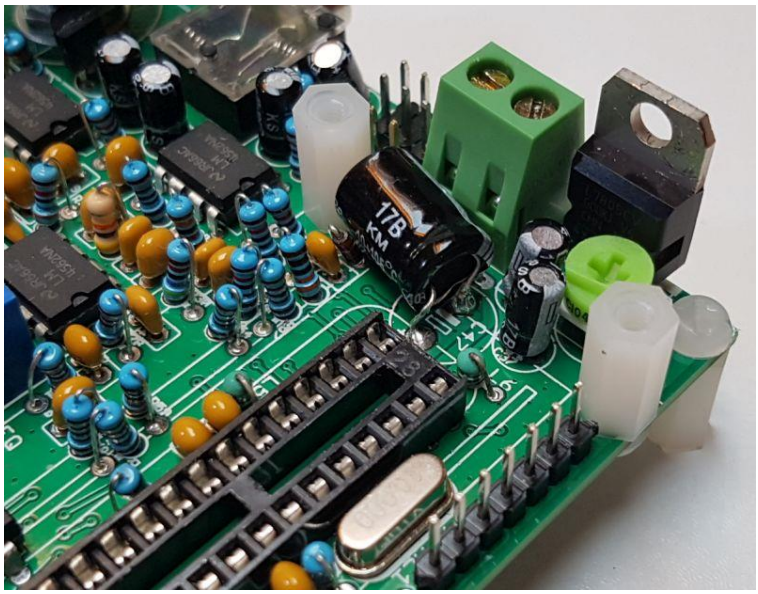
There are six 10uF capacitors in the kit: C21, C22, C24, C37, C38 and C51. These are polarised electrolytic capacitors and MUST be installed with the correct orientation! The capacitor NEGATIVE wire must be installed in the hole indicated on the PCB silkscreen and the layout diagram by the solid black bar; the POSITIVE wire must be installed in the hole indicated on the PCB silkscreen and the layout diagram by the hollow bar and the + symbol.

Electrolytic capacitors are also supplied with one wire longer than the other. The + wire is the longer wire, the – wire is the shorter one (see photo, right).



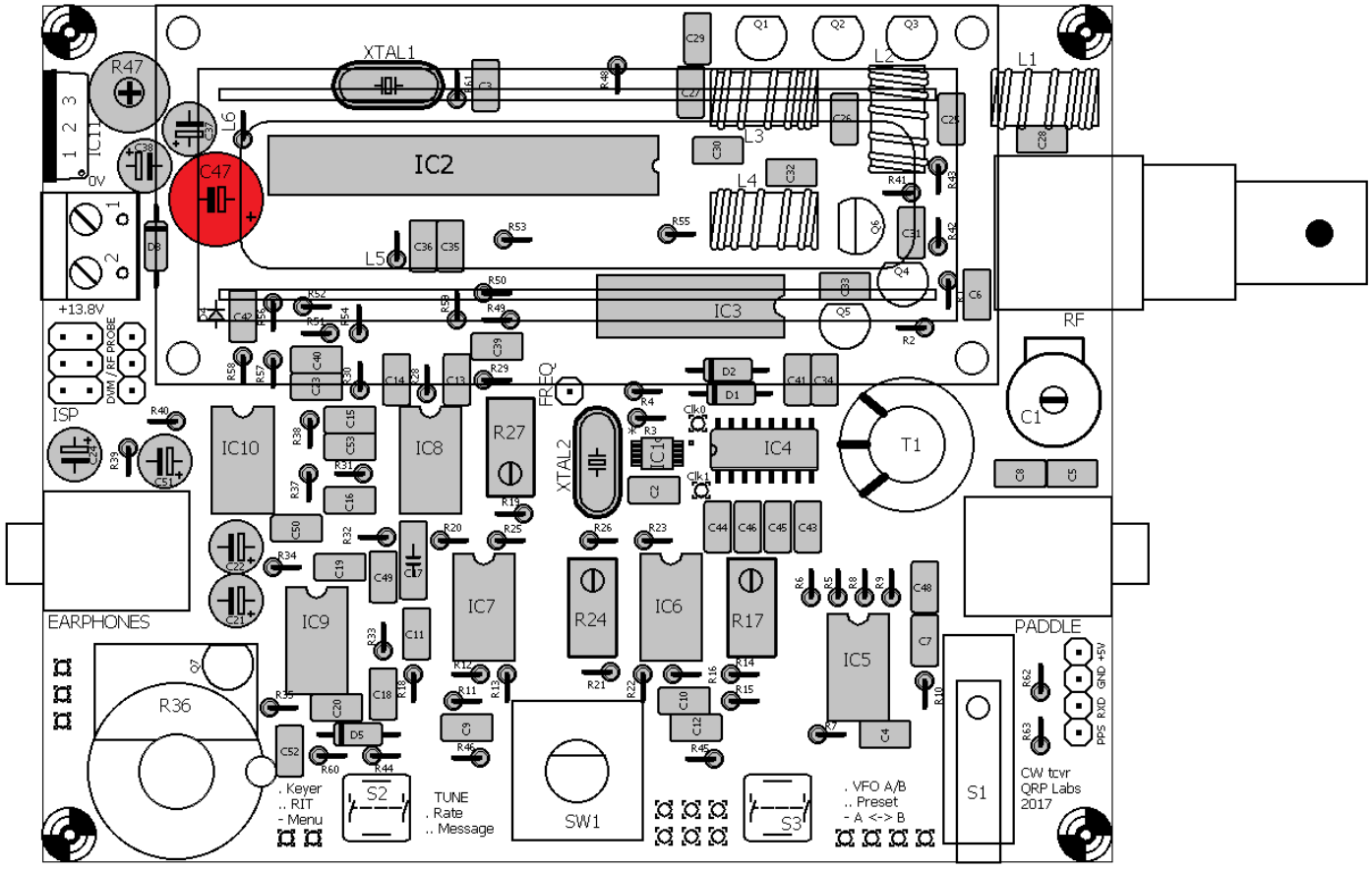
3.41 Install 220uF capacitor C47 (may be supplied as 470uF)

This capacitor is also a polarised electrolytic and must be orientated correctly (see former section).



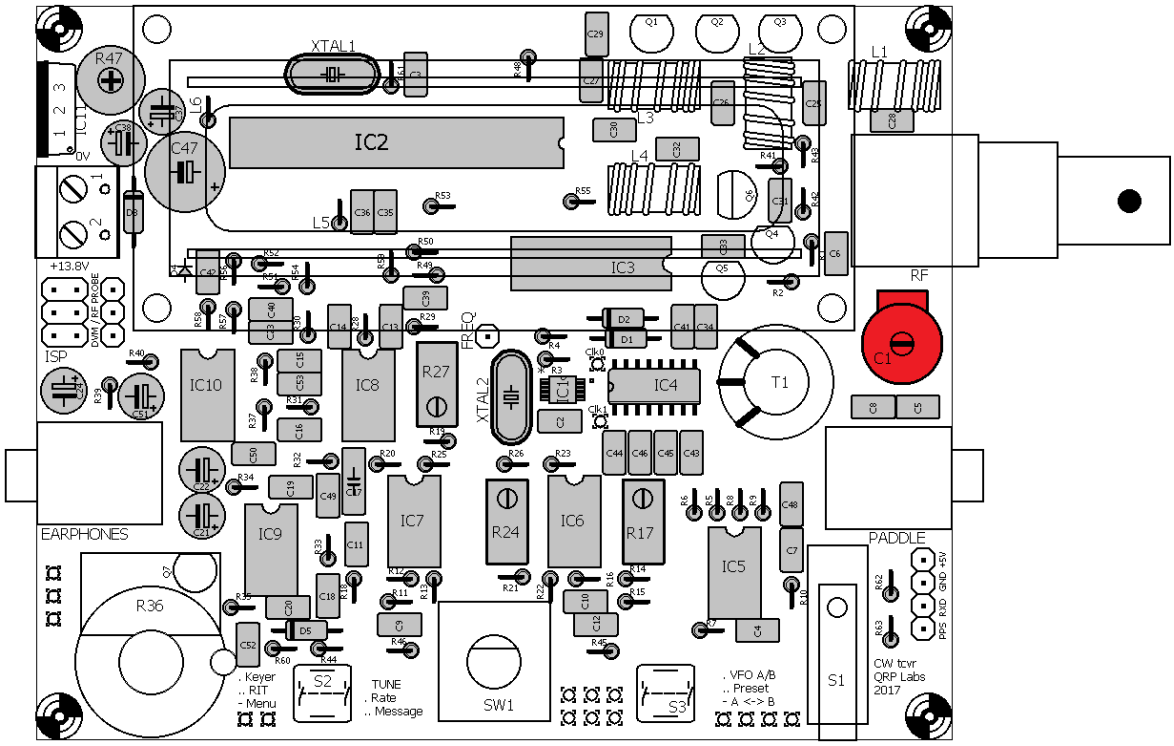
The height dimension of the supplied capacitor in the first batch (up to serial number 500) is enough that in some cases, if installed vertically on the PCB, it may touch some components on the bottom of the LCD module, once the LCD module is installed. Therefore, the recommended way to install this capacitor is orientated horizontally. Insert the wires into the holes (carefully putting the correctly polarised wires in the right holes) then bend the wires so that the capacitor lays horizontally over diode D3 (1N5819). Refer to the photograph (right). When the capacitor is in a suitable position, solder the wires as usual.

Kits from subsequent batches (approximately serial number 500 and above) do not need to mount the capacitor horizontally, because the supplied capacitor is small enough to fit under the LCD.



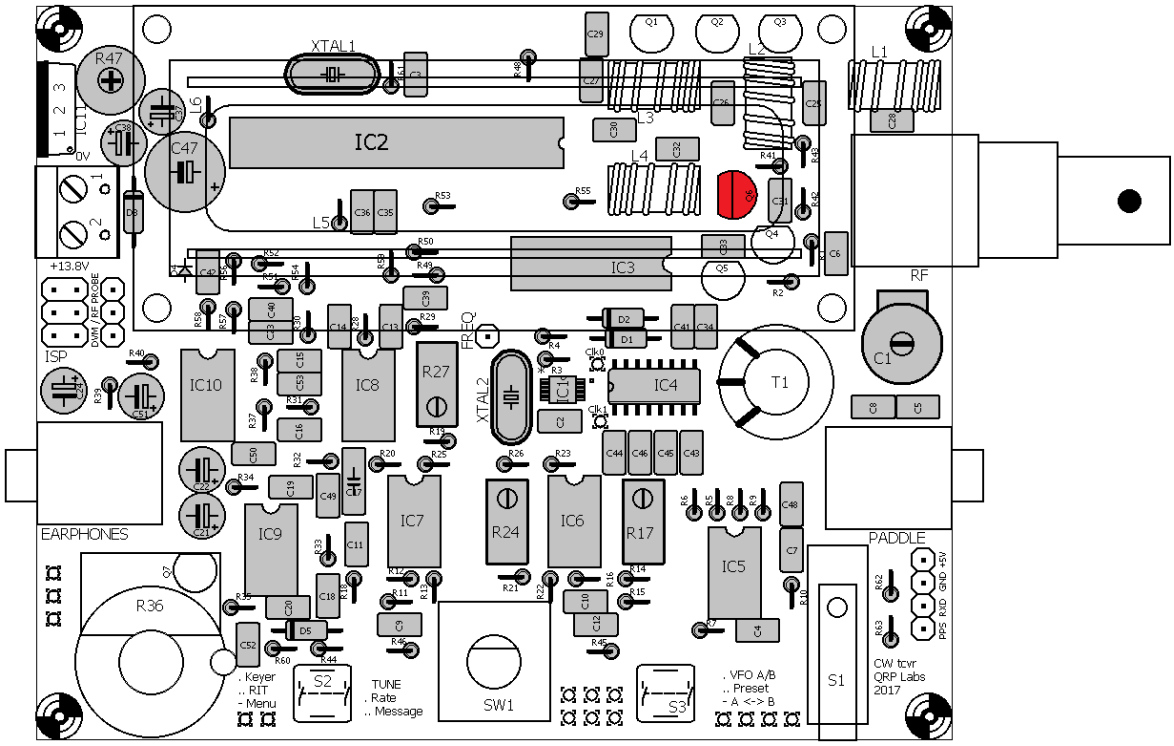
3.42 Install 30pF trimmer capacitor C1

Insert the component pins carefully and solder them QUICKLY to avoid overheating the capacitor which could melt the plastic insulating spacers.



3.43 Install MPS751 transistor Q6 (MPS2907 in first kit batch)

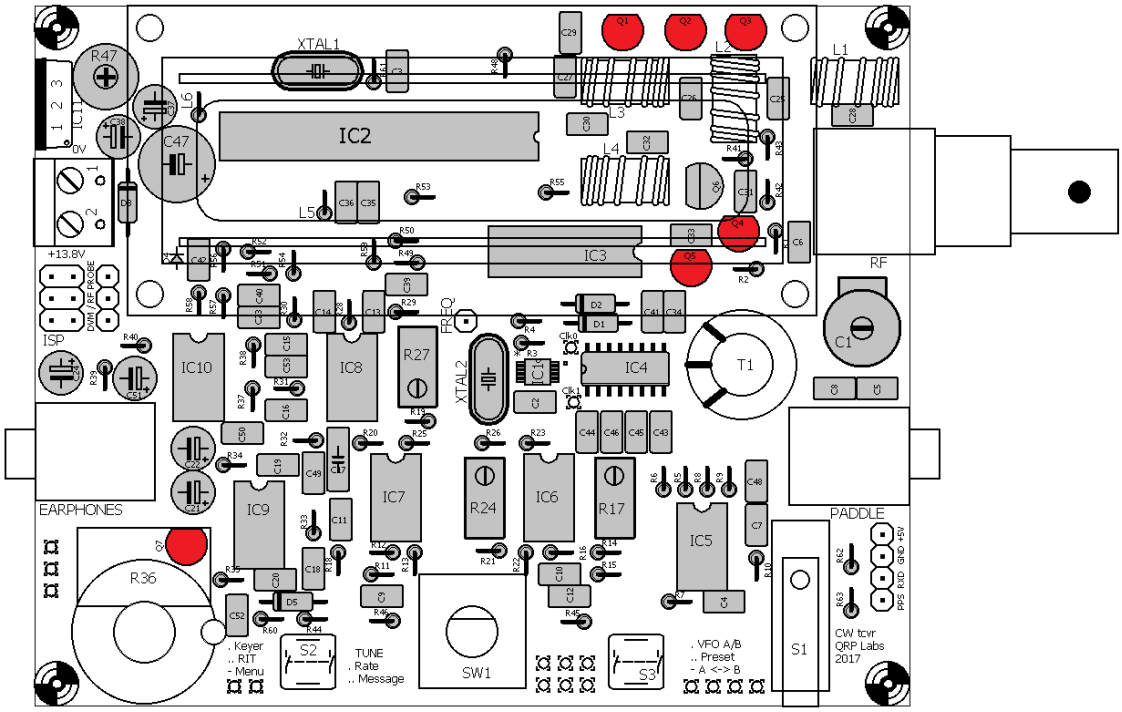
Be careful to correctly identify this transistor by its markings, as the package style is similar to the other transistors. Carefully bend and insert the wires so that the transistor orientation matches the PCB silkscreen and the layout diagram. Carefully push the transistor down towards the PCB until the wires between the PCB and the transistor body are about 5mm long, then solder.



3.44 Install all BS170 transistors

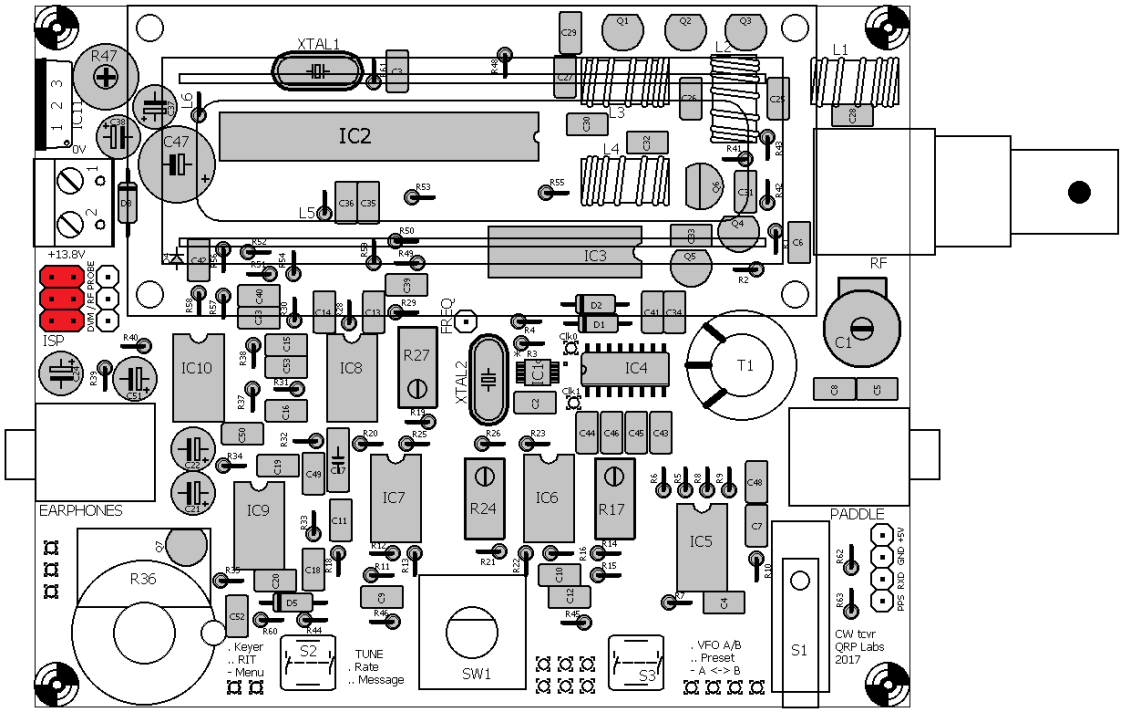
The remaining transistors in the kit are BS170 MOSFETs and there are six of them: Q1, Q2, Q3, Q4, Q5 and Q7. Follow the same installation procedure as the previous section.

Note that Q7 sits under the volume (gain) knob. It is important to ensure that Q7 is pushed down as far as possible towards the board, so that the transistor lead length is only ~4mm, to avoid the top of the transistor blocking the knob.



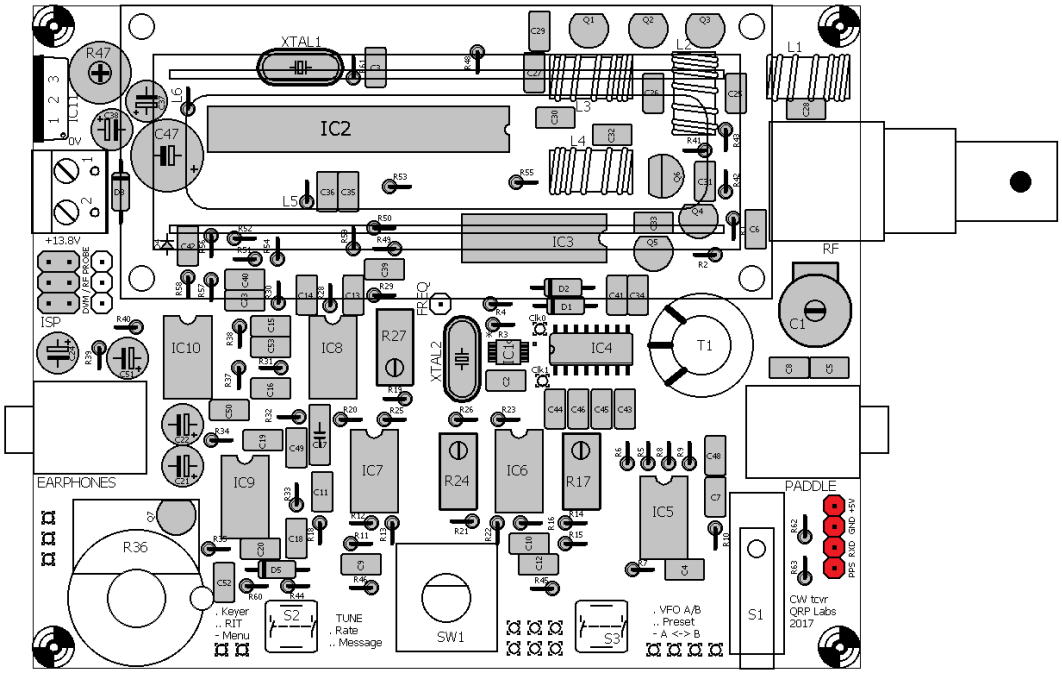
3.45 Install 2x3-pin in-circuit programming header

This pin header can be used to connect an AVR Programmer to apply firmware updates if desired. Insert the SHORT end of the pins into the PCB. Solder one pin in place first and check that the header is nicely seated on the PCB before soldering the other 5.



3.46 Install 4-pin GPS header

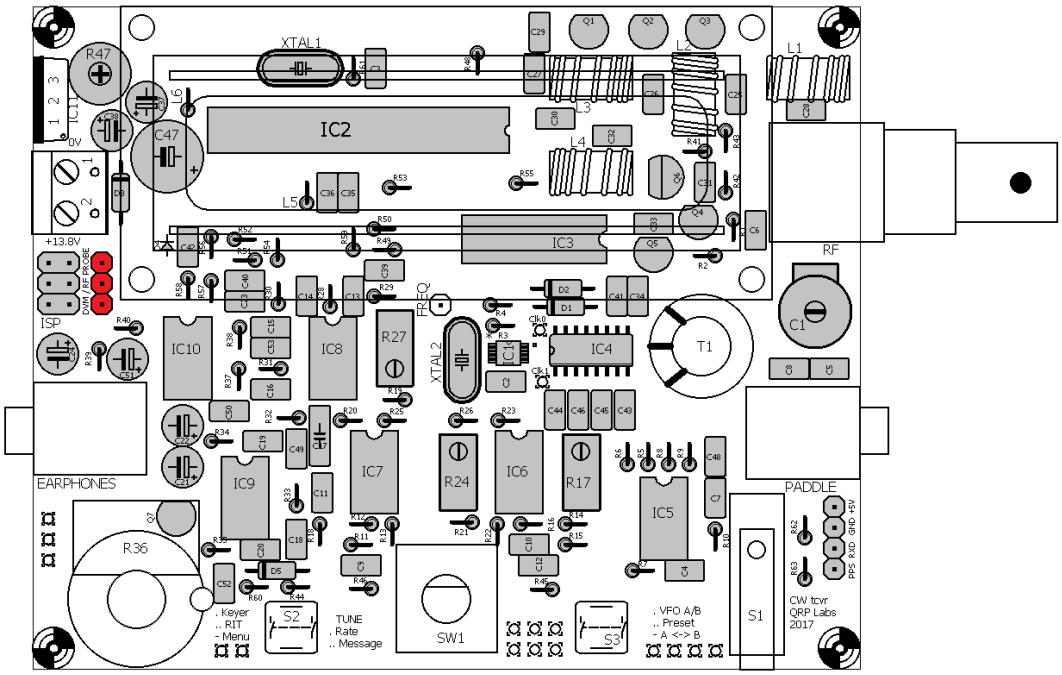
Install a 4-pin header from the Low Pass Filter kit bag, in the position shown. This is used to connect a GPS module if required. Insert the SHORT end of the pins into the PCB.



3.47 Install 3-pin DVM/RF detector header

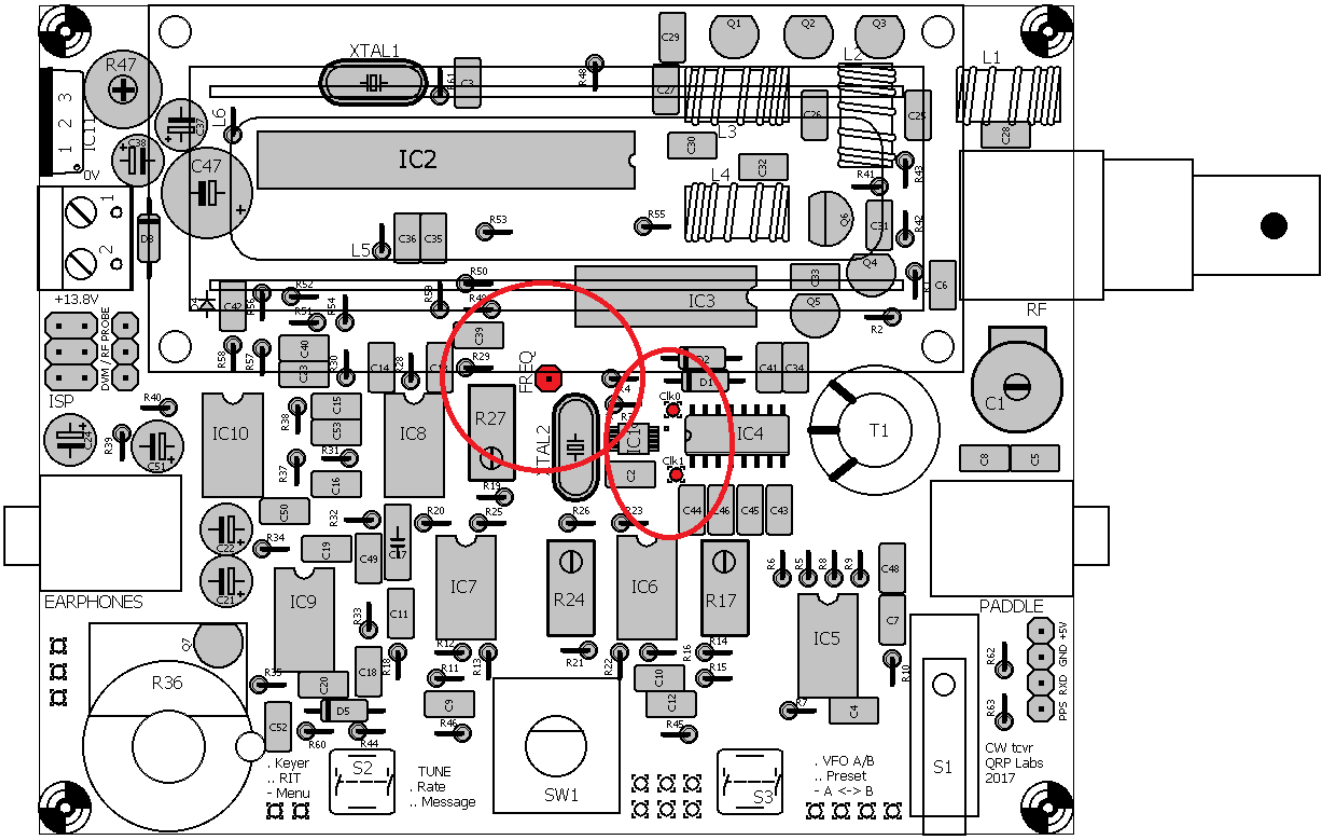
Take the other 4-pin header from the Low Pass Filter kit and carefully cut off 1 pin with a wire-cutter, without damaging its plastic base. The headers are designed to break apart easily like this. Keep the 1-pin section safely.

Install the 3-pin header section as shown. Insert the SHORT end of the pins into the PCB. This is used as the input to the DVM (voltmeter) and RF power meter functions of the kit. You can later jumper pins 1 to 3 to measure the supply voltage with an icon on the LCD during operation.



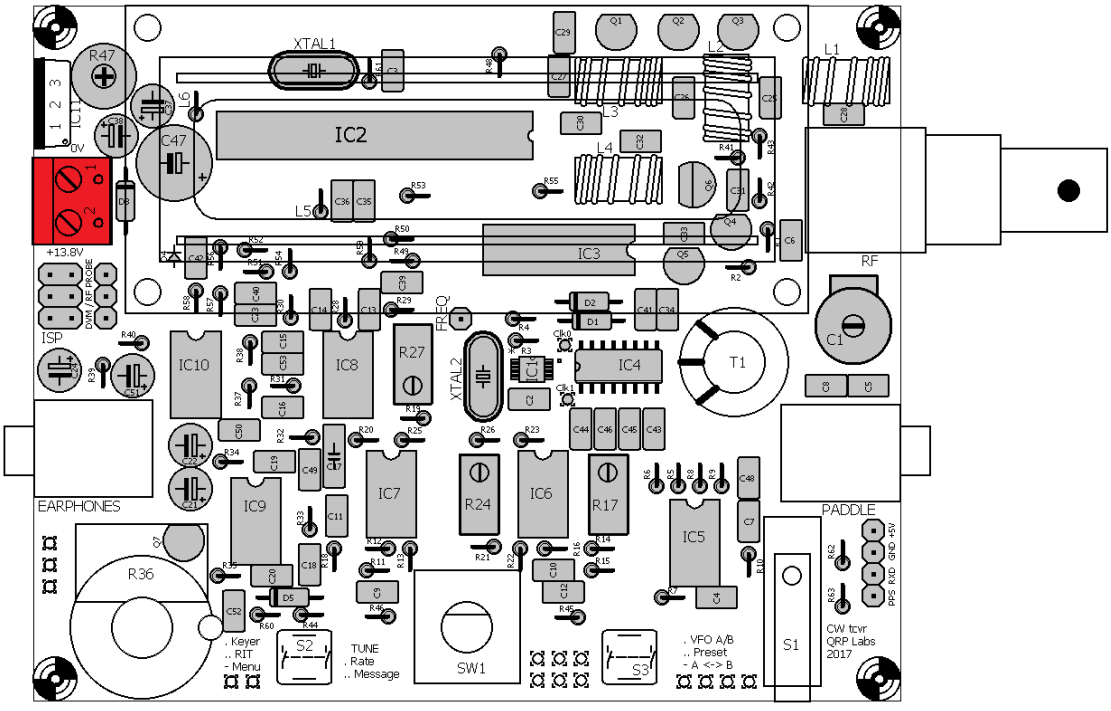
3.48 Install 1-pin test points

Install 1-pin headers as test points, as shown. One of these is taken from the 4-pin header from the Low Pass Filter kit (see previous section); the other two are to be cut carefully from the 16-pin header supplied, leaving the 14-pin section intact.



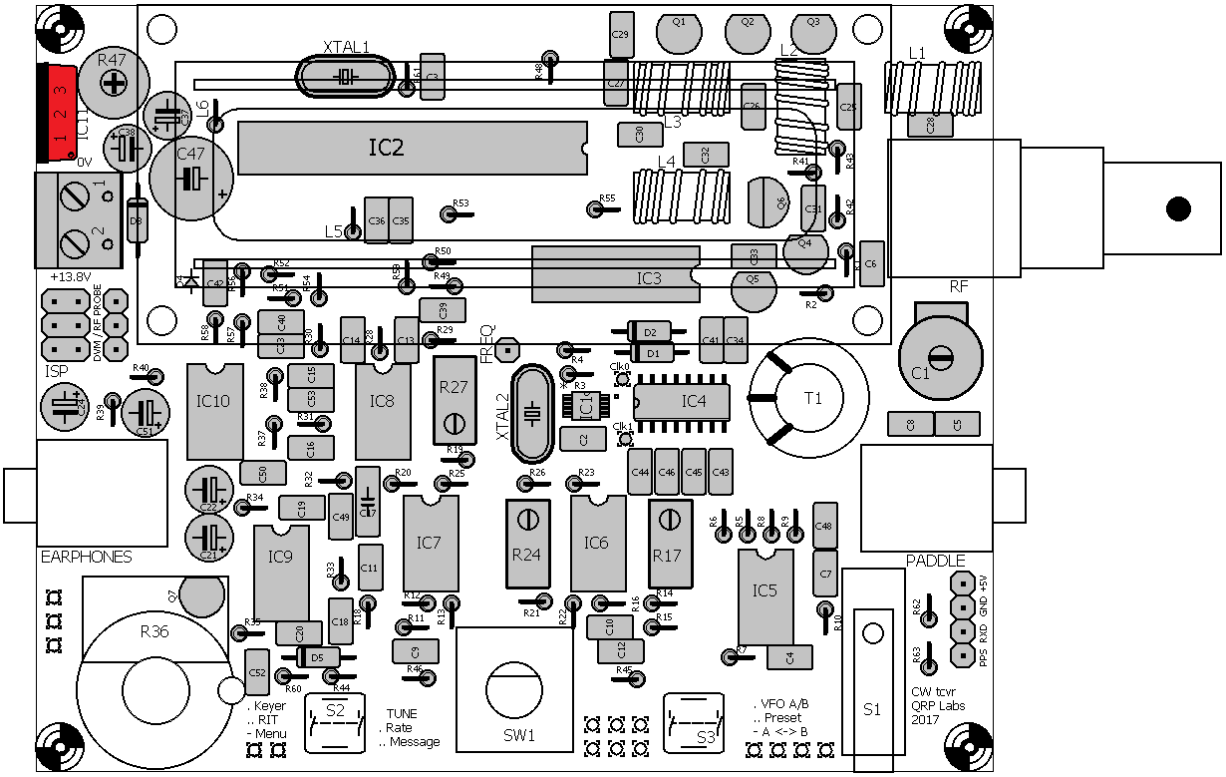
3.49 Install power connector

Install the 2-way power connector as shown, orientated to match the PCB silkscreen and have the terminal openings facing outwards from the PCB to facilitate installing power wires later.



3.50 Install 7805 voltage regulator IC11

Install the 7805 voltage regulator, orientating it so that the metal tab is facing outwards from the PCB. Push the wires all the way down into the holes until the body of the regulator is about 5mm off the PCB surface. Solder and trim the excess wires under the board.

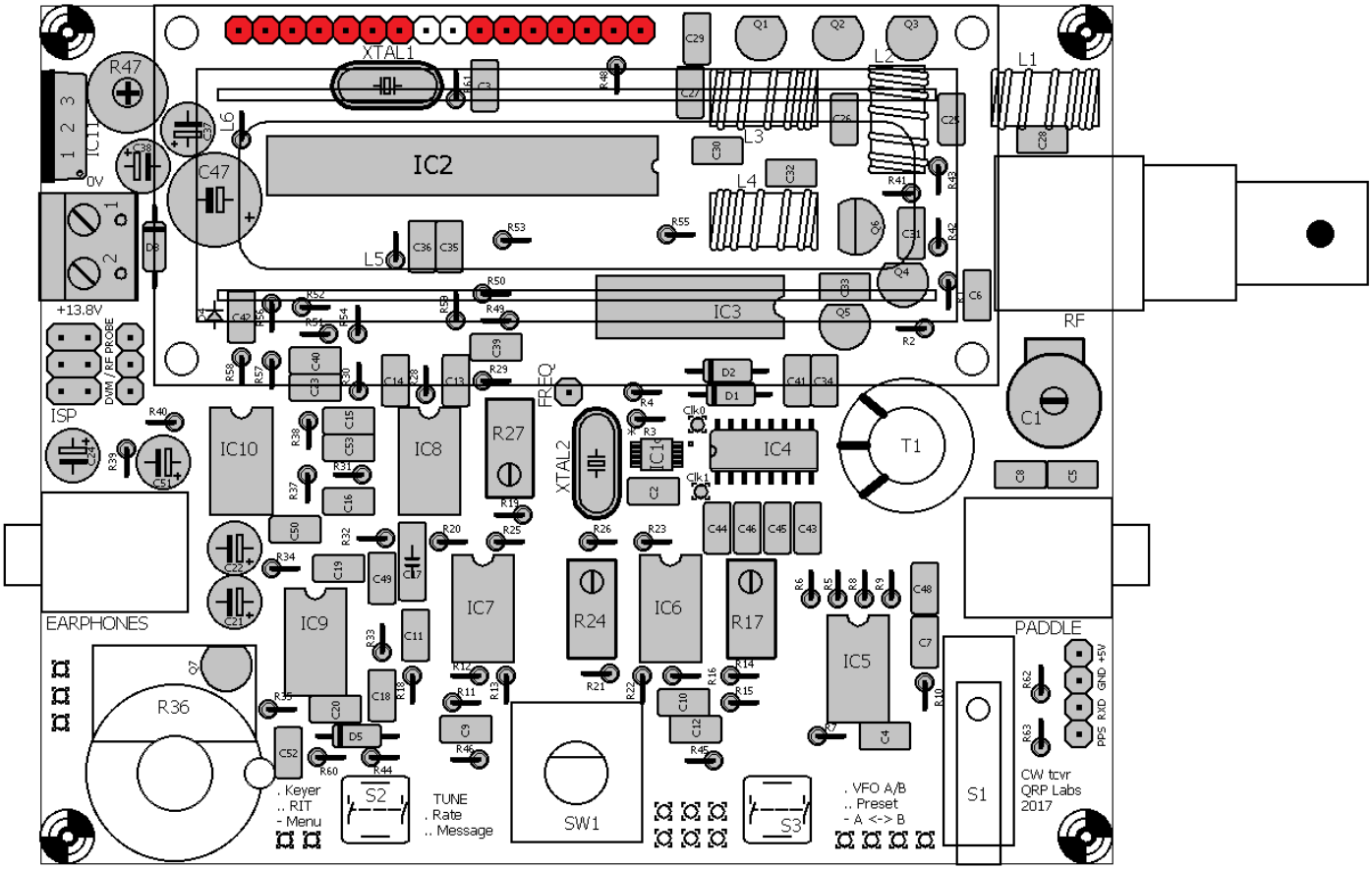


3.51 Install pin headers for the LCD module

The LCD has a 16-way connector. If you have been following me so far, you have already carefully cut off two of the pins, to use in the Clk0 and Clk1 test points. Therefore, you only have a 14-pin header remaining. This is fine, since not all of the pins of the LCD module are used! The LCD module has an 8-bit data bus, but can also be operated in 4-bit data mode, which is what is done in this kit. The 4-bit mode has the advantage that less pins of the microcontroller are used! This is very important in this design, because we require so many control pins from the microcontroller for the various functions of the radio. So, the LCD has 16 pins but the central 4 pins are not used.

Now break the 14-pin header carefully in half so that you have two 7-pin sections. (If you make a mistake and you have one 8-pin section and one 6-pin section, that is still Ok).

Install the two sections into the 16-pin holes on the PCB leaving an empty gap of two pins in the middle. Insert the SHORT end of the pins into the PCB. **Be very careful to ensure that the pin headers are accurately aligned to the PCB, so that later when we install the LCD module everything fits accurately.**



3.52 Wind and install toroid L4

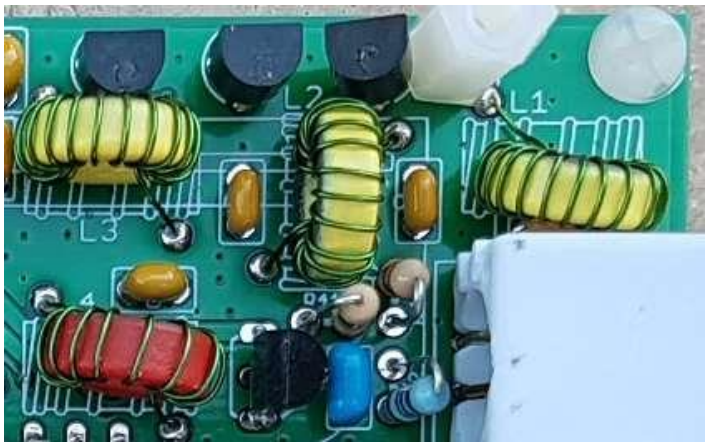
L4 is type T37-2. It is a small ring with red paint on one side. Each time the wire passes through the hole in the middle of the toroid, this counts as one turn. The number of turns depends on the band of your kit, refer to the following table. Inductance values are approximate and will depend on variations in the core, and how tightly you wind the turns. Do not worry about these variations which are not critical in this case.

Band	Value	Turns
80m	2.3uH	24
60m	2.3uH	24
40m	1.0uH	16
30m	0.78uH	14
20m	0.40uH	10
17m	0.32uH	9

Try to keep the wire quite tight (but not so tight that you break the wire). Try to spread the turns evenly around the toroid. Leave about 2cm or 3cm of wire at the ends.

The wire is coated with an enamel insulating paint and it is **CRITICAL to remove this enamel at the soldered joints otherwise there will be no electrical connection through the toroid! This is the number 1 cause of problems with QRP Labs kit construction: failure to remove the wire enamel.**

One method of removing the wire enamel is to scrape it off at the ends, either with sandpaper, or carefully scratching with a knife or wire cutters. However, my favourite method is just to burn off the enamel. For quite a number of years, the type of enamel used on copper wire can be burnt off using the temperature available from an ordinary soldering iron, immersing the wire in a blob of solder. (This was not the case, with much older wire found in vintage valve/tube equipment). You could also use a cigarette lighter to burn off the enamel.

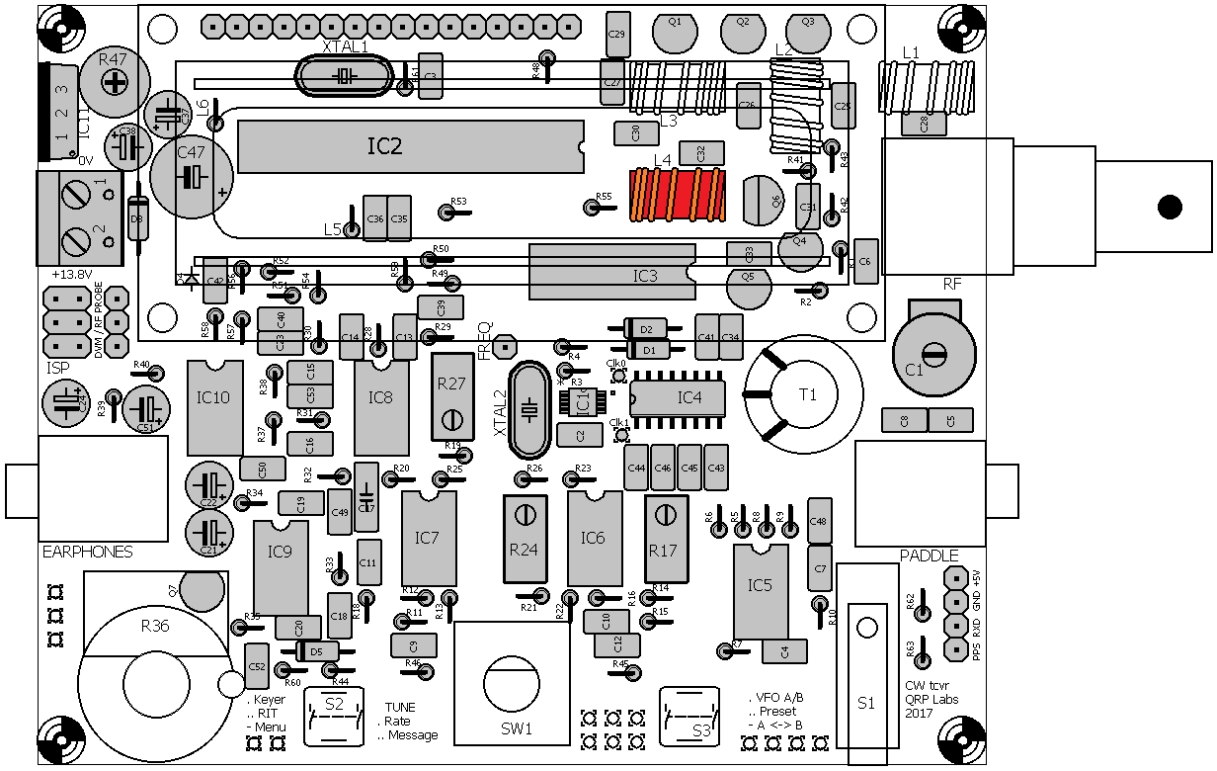


Insert the ends of the wires into the correct holes of the PCB and pull it through tight, so that the toroid sits up vertically on the PCB. See photo, right.

Bend the wires over so that the toroid does not fall out while you are trying to solder the wires. Cut off the excess wire using wirecutters, leaving only about 2mm protruding through the PCB on the lower side. Now apply solder quite generously from the soldering iron. Hold the soldering iron to the joint for a few seconds – I usually count to 10 slowly – and the solder will surround the wire, which will become hot enough to burn off the enamel. You can sometimes see a small puff of smoke when the enamel burns off.

Carefully inspect the soldered joints with a magnifying glass to make sure that the wire is correctly soldered. If it looks like the solder hasn't flowed nicely and adhered to the wire, then this is usually a sign that the enamel probably hasn't been burnt off.

If you have a DVM it is a good idea to check for DC electrical continuity (zero ohms resistance) between the two ends of the wire. If you do NOT have a DVM and if the radio doesn't work, then we can use the built-in test equipment later, to trace the fault.

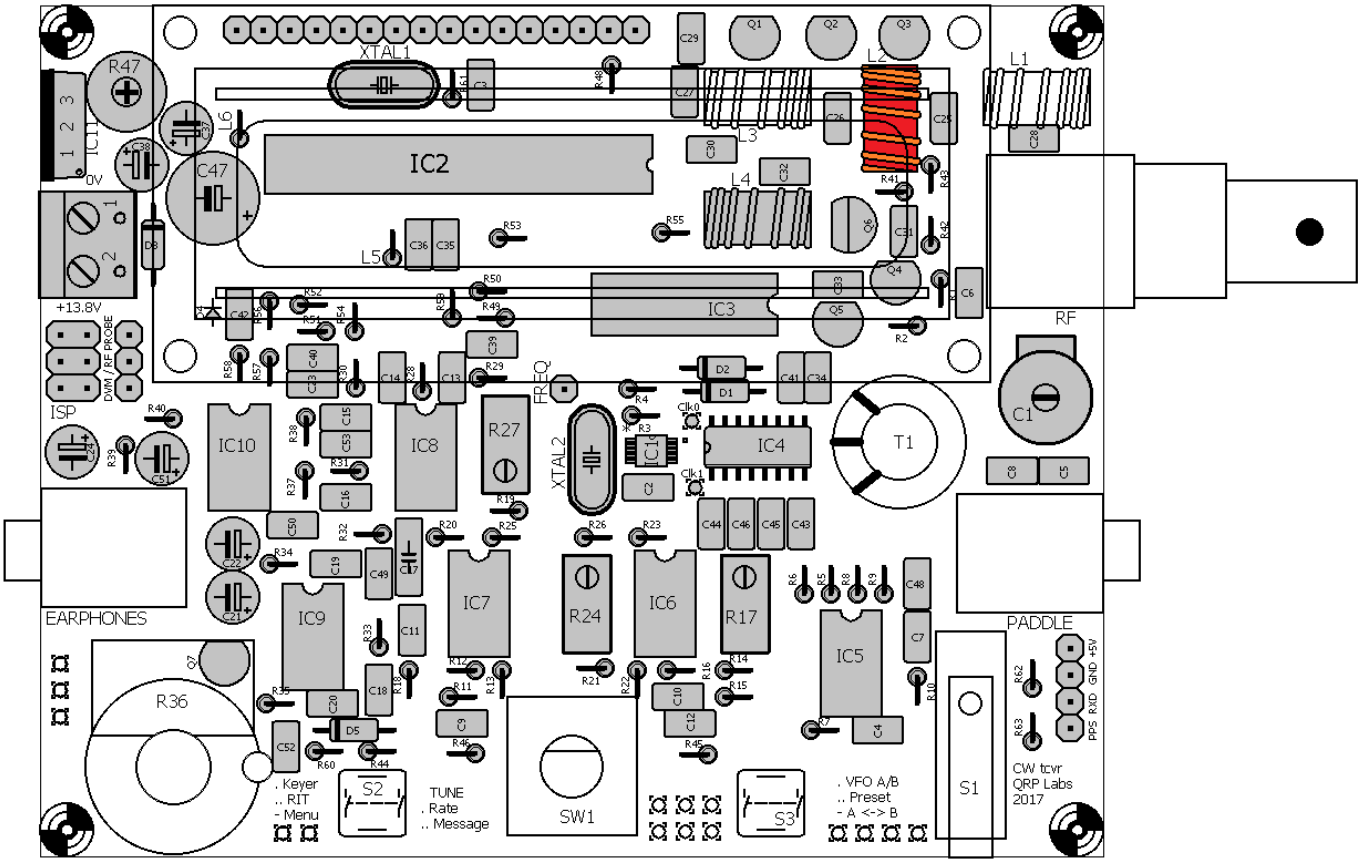


3.53 Wind and install toroid L2

L2 is a small toroid ring painted yellow or red on one side. It is part of the supplied Low Pass Filter kit bag. Again, keep the wire tight and evenly spaced around the toroid. Installation of the inductor is similar to the previous section. **Remember to remove the wire enamel and check!**

The number of turns is band-dependent, refer to the following table. Inductance values are approximate and will depend on variations in the core, and how tightly you wind the turns. Do not worry about these variations which are not critical in this case.

Band	Toroid	Colour	Value	Turns
80m	T37-2	Red	3.0uH	27
60m	T37-2	Red	2.3uH	24
40m	T37-6	Yellow	1.7uH	24
30m	T37-6	Yellow	1.3uH	20
20m	T37-6	Yellow	0.90uH	17
17m	T37-6	Yellow	0.67uH	15

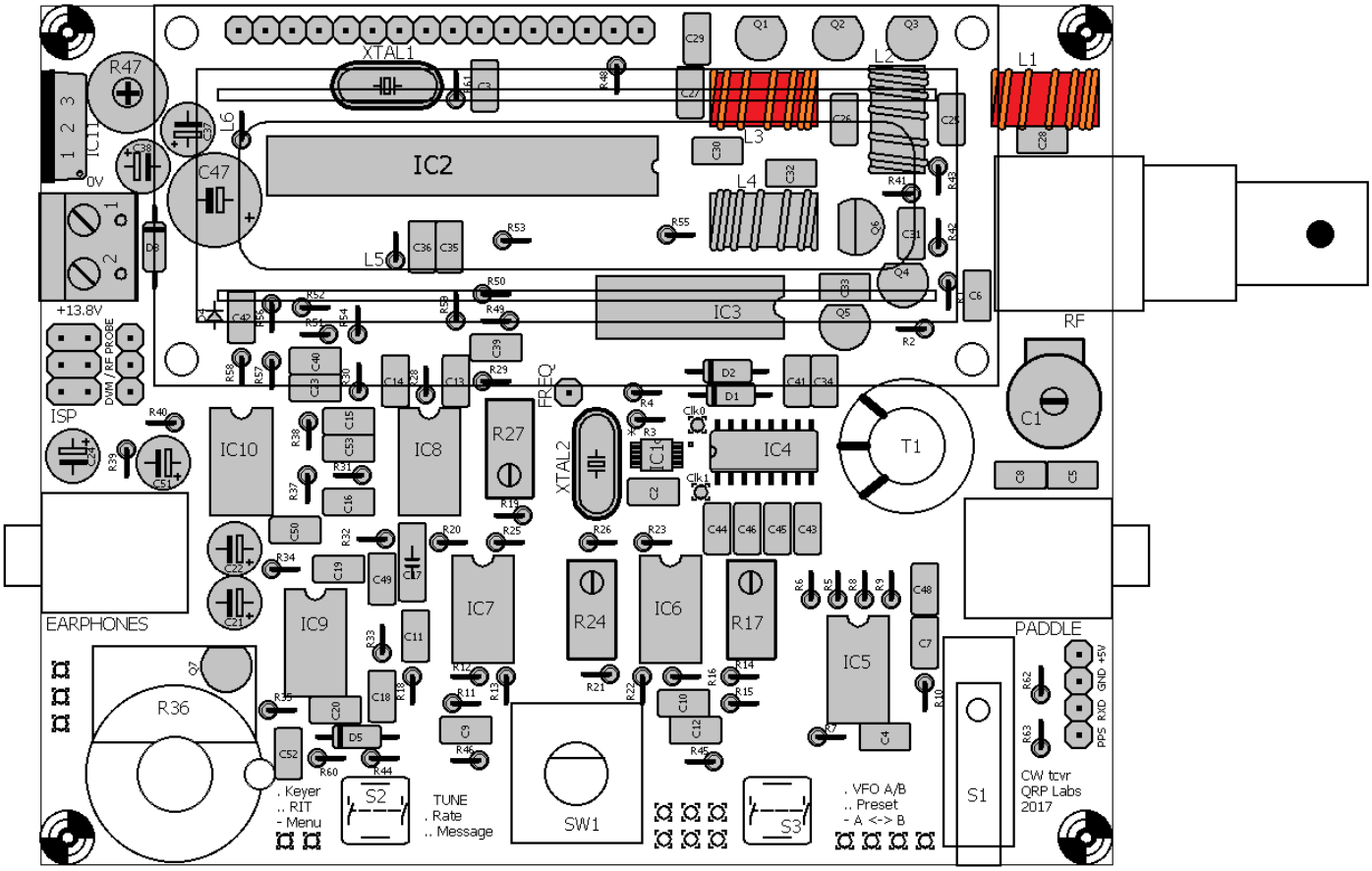


3.54 Wind and install toroids L1 and L3

L1 and L3 are small toroid rings painted yellow or red on one side. They are part of the supplied Low Pass Filter kit bag. Again, keep the wire tight and evenly spaced around the toroid. Installation of the inductor is similar to the previous section. **Remember to remove the wire enamel and check!**

The number of turns is band-dependent, refer to the following table. Inductance values are approximate and will depend on variations in the core, and how tightly you wind the turns. Do not worry about these variations which are not critical in this case.

Band	Toroid	Colour	Value	Turns
80m	T37-2	Red	2.4uH	25
60m	T37-2	Red	2.1uH	23
40m	T37-6	Yellow	1.4uH	21
30m	T37-6	Yellow	1.1uH	19
20m	T37-6	Yellow	0.77uH	16
17m	T37-6	Yellow	0.55uH	13



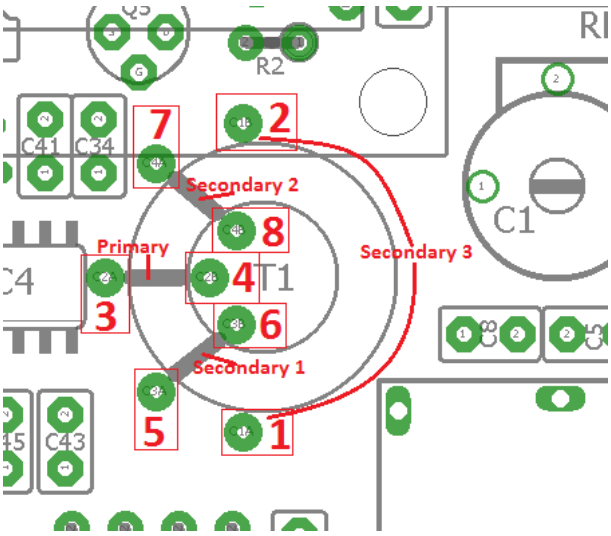
3.55 Wind and install transformer T1

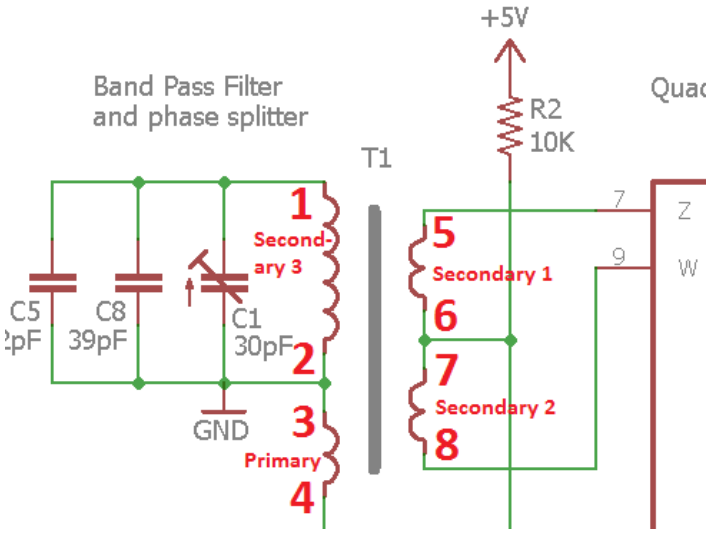
Now we come to the only really tricky piece of the assembly, which is the receiver input transformer T1. Follow these instructions carefully, it is tricky but quite feasible if you go step by step.

In the end, you are going to end up with an installed transformer which hopefully looks something like the photo (right, shows 20m version).

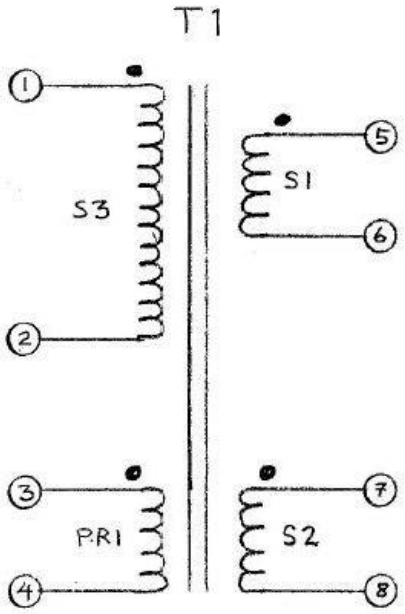
This transformer has FOUR windings. Three identical short windings, and one long winding. There are therefore eight wire ends, which must all be soldered into the correct holes on the PCB, and with the enamel properly removed.

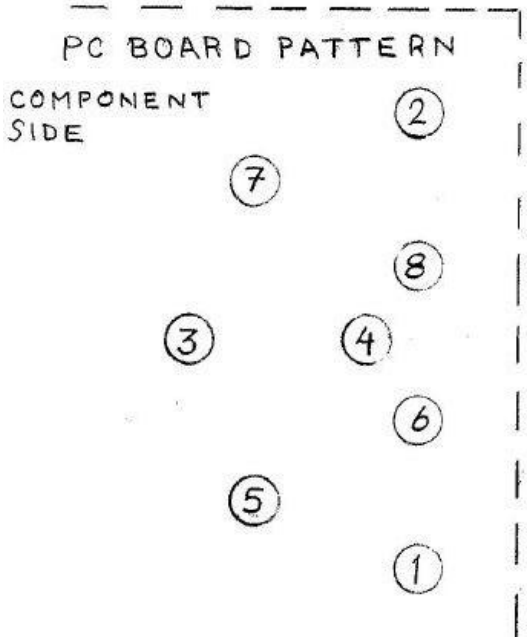
Here are two diagrams which enumerate each of the windings, wire endings and holes on the PCB, on both the layout diagram and on the circuit (schematic) diagram. It should help to explain diagrammatically which wires must go where.

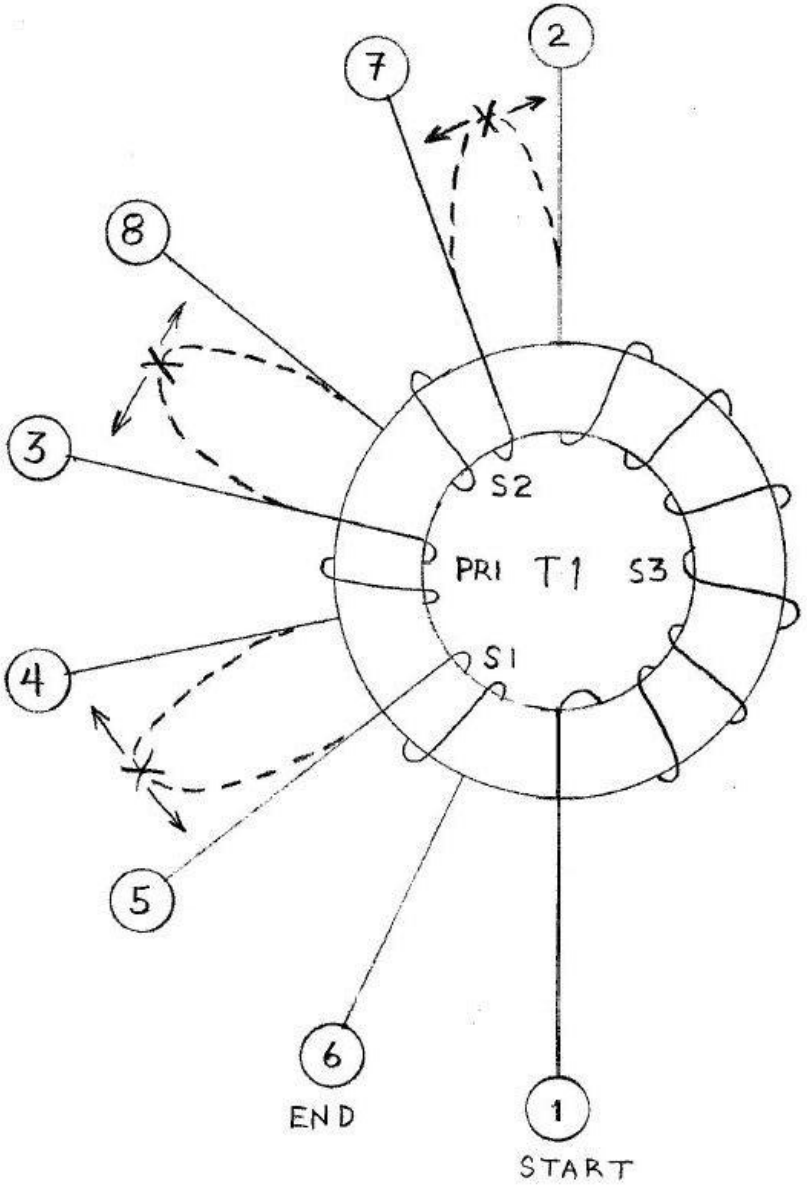




On the following page is a diagram hand-drawn by Ed WA4MZS (thanks Ed!) which may also clarify the construction and installation of T1.



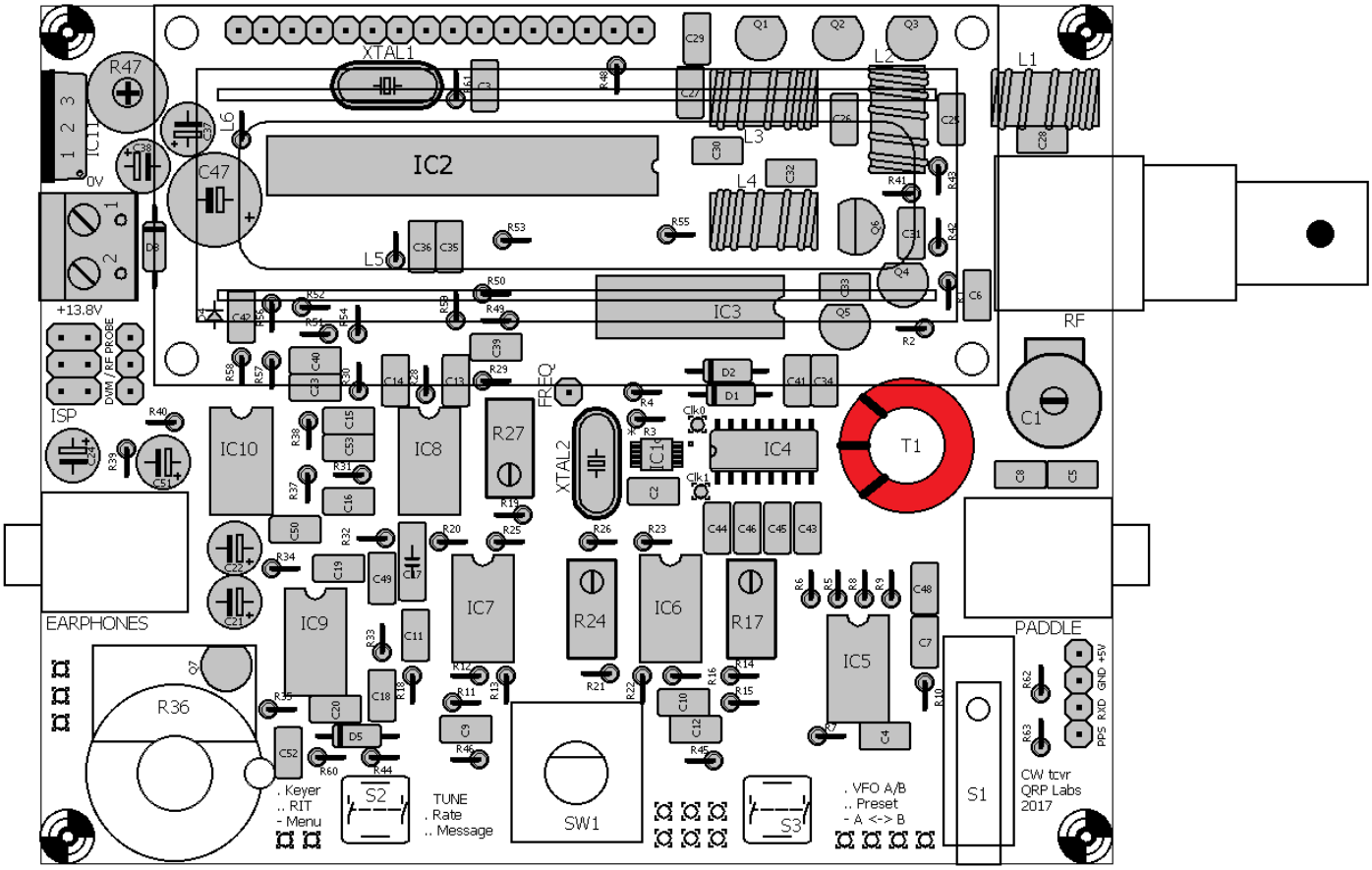




BAND	PRI	S1	S2	S3
80m	5T	5T	5T	68T
60m	5T	5T	5T	46T
40m	5T	5T	5T	38T
30m	4T	4T	4T	30T
20m	3T	3T	3T	30T
17m	3T	3T	3T	22T

CORE = T50-2

WIND W/ # 28 ENAM.
(0.33 mm)



The four windings on T1 must all be in the same “sense”. There are two ways to wind toroids. You might call them left-handed and right-handed; clockwise and counter-clockwise; whether the wire goes through the toroid from top to bottom, or from bottom to top. Whatever you call it, all the four windings have to be the same, to be sure to get the phasing to the quadrature sampling detector correct.

The number of turns in each winding depends on the band you are building the kit for. Refer to the following table. For convenience, the remainder of the instructions in this section refer to the 20m version (30 + 3 + 3 + 3 turns).

But MAKE SURE you wind the correct number of turns for your band!

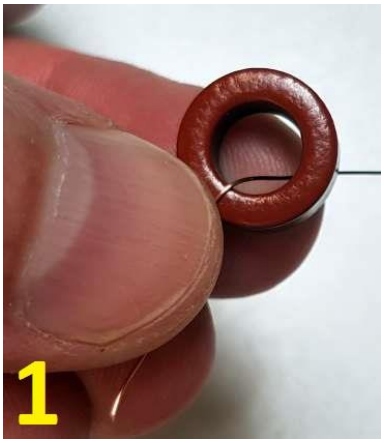
Band	Primary	Secondary 1	Secondary 2	Secondary 3
80m	5	5	5	68
60m	5	5	5	46
40m	5	5	5	38
30m	4	4	4	30
20m	3	3	3	30
17m	3	3	3	22

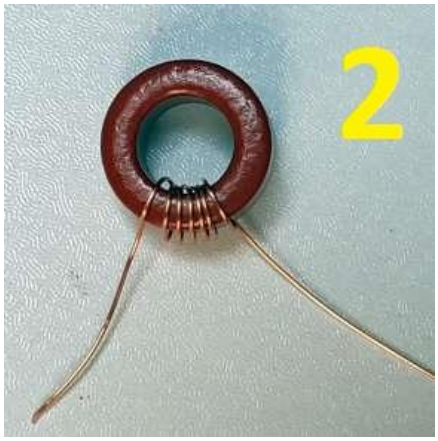
Note for 80m and 60m versions: the toroid ring is not large enough to neatly hold all of those turns in a single flat tidy winding. The large secondary winding WILL end up looking messy, with overlaps in some places. You should try to ensure any overlaps turns are evenly spaced throughout the winding! Do not try to wind one neat layer, then wind the remaining turns as an additional neat layer on top of that. Just go with the messy overlapping turns and don’t worry about it: everything will work fine, regardless.

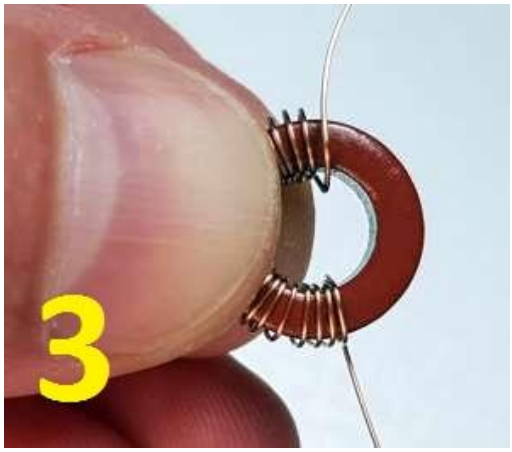
In all cases, there is one long secondary winding, and three other short identical windings. To make things easier, we will wind all windings together in one go, this will guarantee that the "sense" of each winding will be the same. At the intended breaks between the windings, we will leave large loops of wire, that we can later cut one by one to make sure the wires go in the correct holes.

There are a lot of steps listed below but in reality, it is easier DOING it than it is writing instructions how to do it. Take it patiently, step by step:

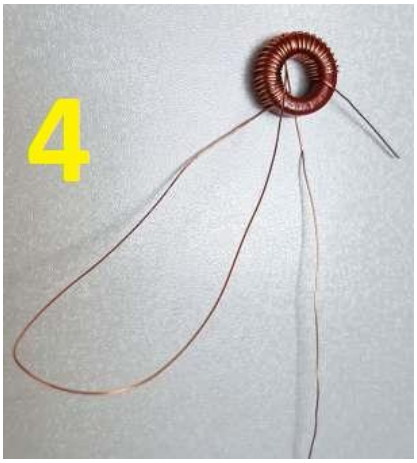
1) Hold the toroid between thumb and finger, and thread the first turn of the wire through from top to bottom. Leave about 3cm of wire at the free end.
2) Apply tension to the wire after each pass through the centre hole, to try to keep the windings tight and even. The wire turns should sit neatly side-by-side on the toroid, without overlapping.
3) Grip the toroid between thumb and finger as you wind.

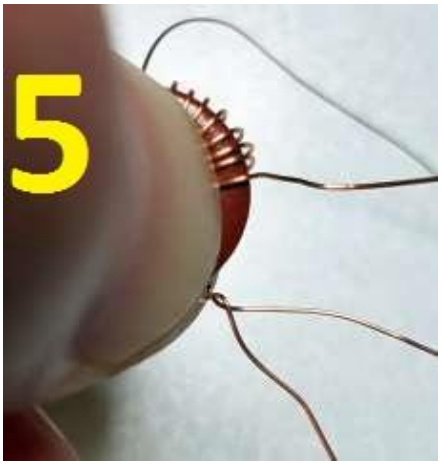


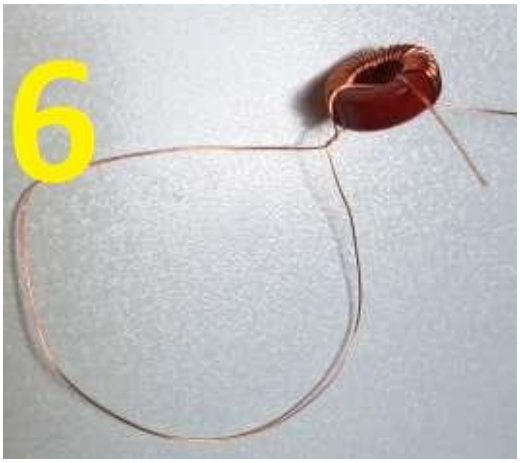




4) When you have completed 30 turns, pass the wire through the hole for the 31st turn but leave a large loop of wire between the 30th and 31st, without pulling it tight.
5) Grip the toroid and wound turns tightly between thumb and finger, and with the other hand apply some twists the wire loop, tight next to the toroid.
6) The result is a wire loop after 30 turns, as shown.





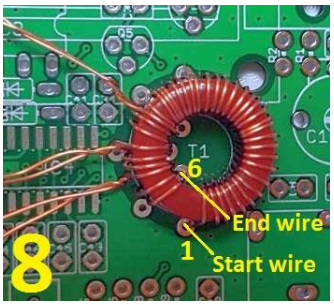


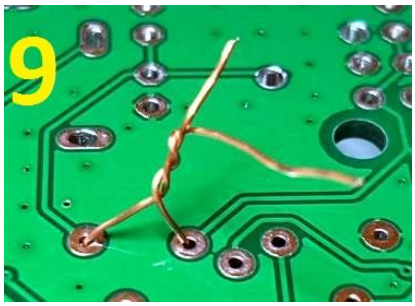
7) Do the same thing for two more loops, which are between the 33rd and 34th turns, and 36th and 37th turns respectively. It is easy to lose count. An easy way is after step 6, pass the wire through the hole 3 more times, and on the 3rd one, make a loop. Similarly count three more and make a loop on the third. Then finally three more turns to complete the

toroid's 39 windings. Count the windings to make sure you have 39. When you are sure everything is fine, cut the wire leaving about 3cm of wire free at the end.

8) **The remaining steps show the transformer installation on an otherwise empty PCB, to make the explanation clear.** Thread the original start of your winding (from step 1) into hole 1 in the diagram. Thread the final end of your winding (from step 7) into hole 6.

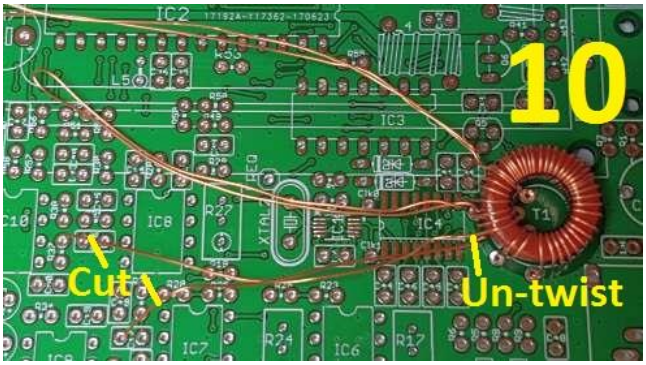
9) Twist these two wires under the board to keep the toroid in place while you deal with the remaining wires.

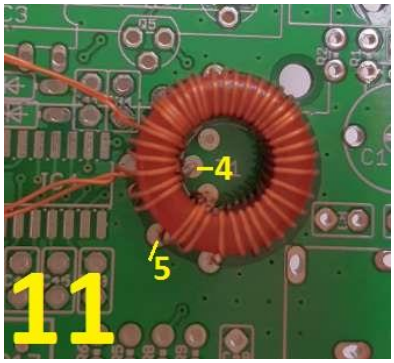




10) Now cut the 3rd loop (the one nearest the end of the winding work), and un-twist the twisted section you made near the toroid body.

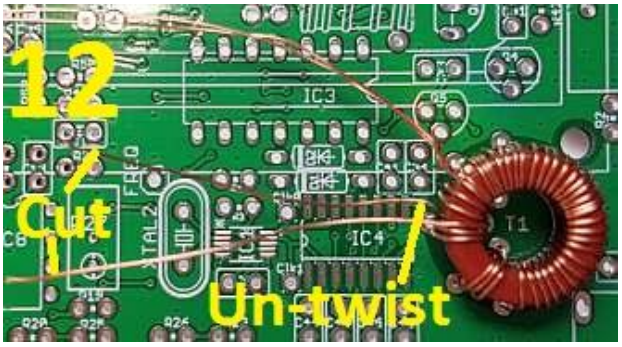
11) When you cut the loop, you therefore created two wires. One of these came over the top of the toroid, and you can easily verify that this is the one which has three turns through the toroid then passes through hole 6. So, insert this wire into hole 5. Once again to be clear: you should now have a winding of three turns (which is labelled "secondary 1", above), with one end inserted into hole 6 and one end inserted into hole 5.
The other wire came from UNDER the toroid. You must push this wire towards the toroid and pull it out through the centre hole of the toroid. Now thread it through hole 4.
Under the board, twist the two new ends of wire together to keep them in place.

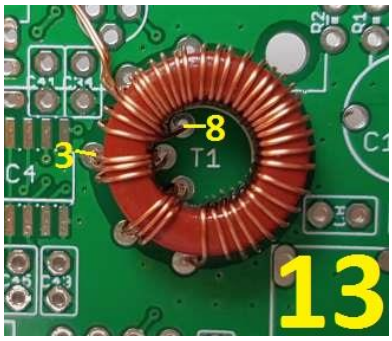




12) Next cut the 2nd loop, and un-twist the twisted section near the toroid body.

13) Similar to step 11; the wire which came over the toroid body has three turns then goes into hole 4, which you did in step 11. Insert this wire-end into hole 3.
The other wire end that came from under the toroid body should be gently pushed toward the toroid, and pulled up through the centre hole of the toroid. Insert that wire end into hole 8. So now you have the "primary" winding consisting of three turns, between holes 3 and 4.
Twist the two new wires under the board again, to keep everything in place.

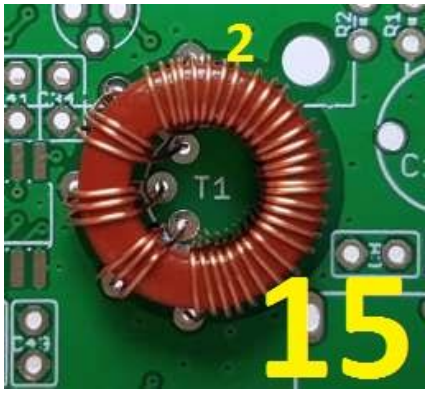


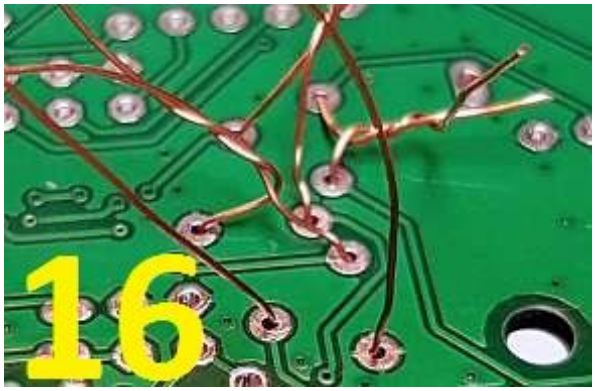


14) Finally, cut and un-twist the loop which you created first, which was between the 30th and 31st turns of the toroid winding. Push the wire which came over the toroid body, into hole 7. Now you have three turns of wire, which make up “secondary 2”, between holes 7 and 8.

15) The last wire came from under the toroid body when you cut the loop; this wire is the other end of the 30 turns “secondary 3” winding, so insert it into hole 2. It is already near to hole 2 and you don’t need to push it under the toroid body as you did in previous steps.
Take a moment now to review the situation. You should be able to identify the four windings of T1, and squeeze them together neatly as in the photo, to verify that each end of the 3-turn windings goes into the correct holes.

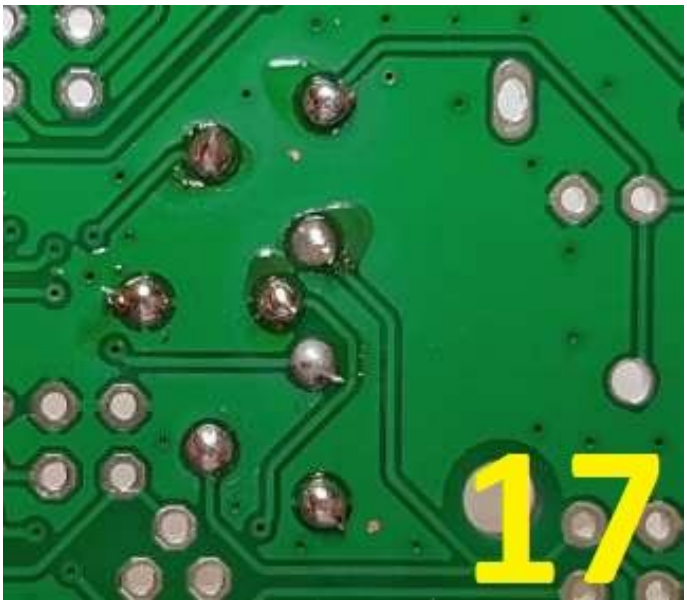
16) Under the PCB, you should have three pairs of twisted wires, and one pair (that you installed last), un-twisted.





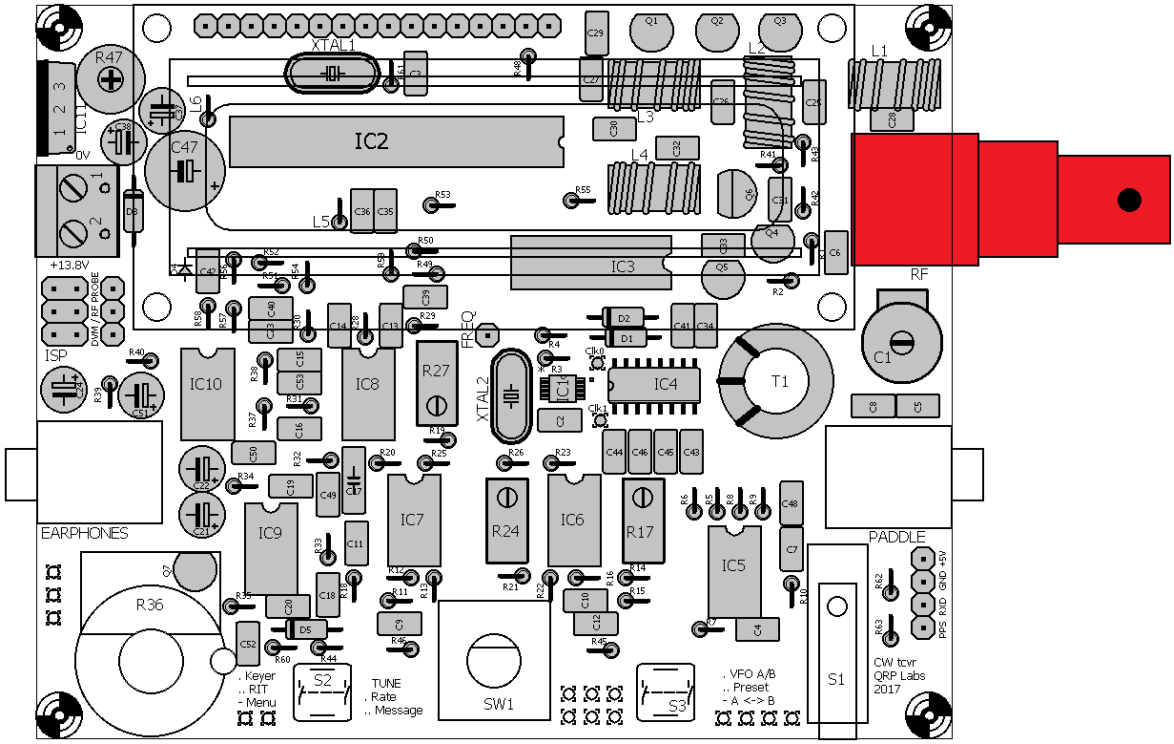
17) Now you can solder the eight connections under the PCB. I recommend doing one pair of wires at a time; this way, the other wires will hold the toroid in place and stop it falling out. Start with the two un-twisted wires. Pull each wire tight, bend it over at about 45-degrees, and cut it 1-2mm away from the PCB surface. Having bent the wire prevents it from falling out. Now solder the wire. Remember to hold the soldering iron to the joint for 10 seconds or so, to allow the enamel insulation to burn off.
Repeat for all the other wires, one pair at a time, until all eight are soldered. If you have a DVM, check for DC continuity (zero ohms resistance) across each winding. If you do not get the expected continuity, then it means either
a) you have not managed to scrape or burn away the enamel insulation properly so there is no electrical connection AND/OR
b) you put the wires into incorrect holes AND/OR
c) your expectation is wrong because you have not identified which pad on the PCB is labelled 1-8 in the diagram.

18) The final picture (18) shows the toroid installation completed.



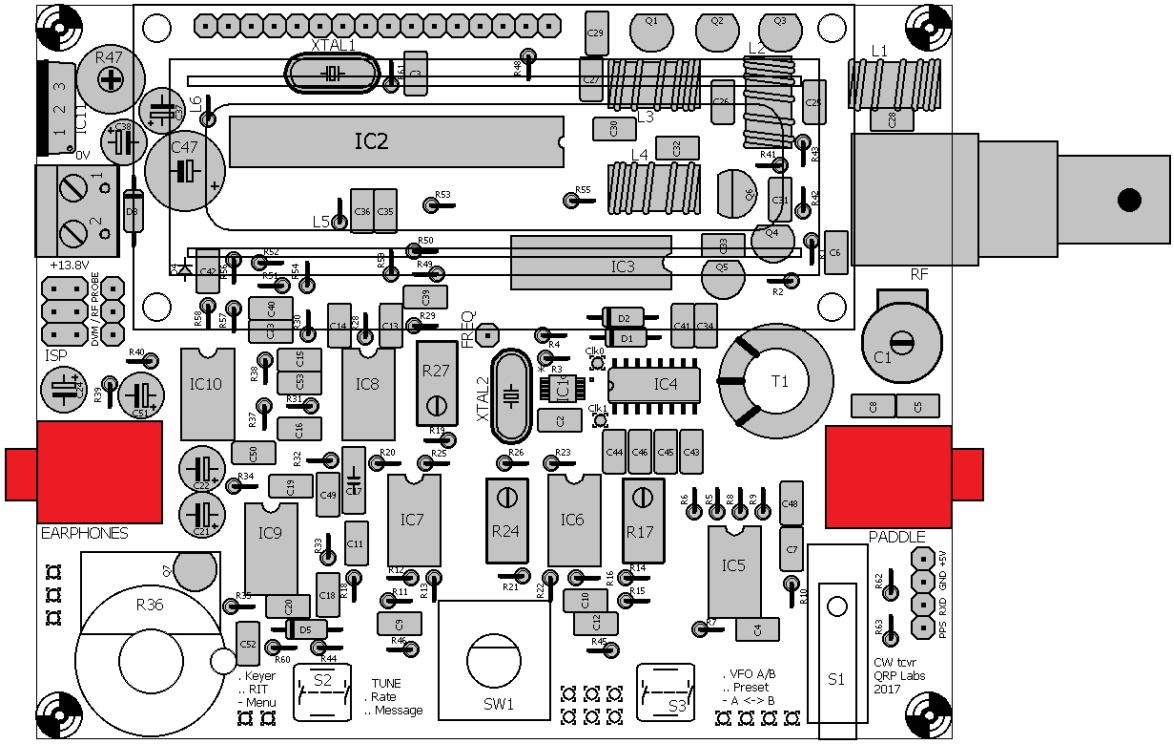
3.56 Install RF output BNC connector

Solder one pin first, to check that the alignment is correct; when happy, solder the remaining pins. The connector has RF and Ground connections, and two fat rods which are for mechanical stability only.



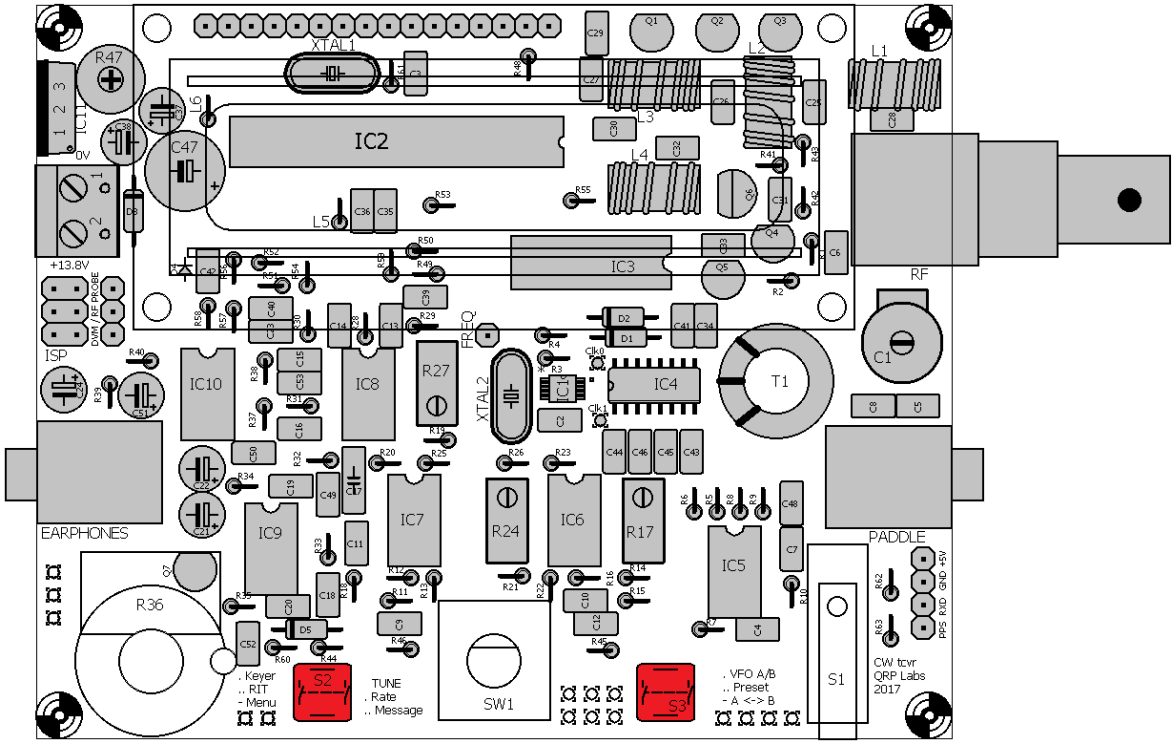
3.57 Install 3.5mm stereo connectors

These connectors are used for the audio output (earphones) and for optional connection of a paddle. The sockets are a VERY tight fit, but they DO fit, with care, eventually. When you are happy that the socket is neatly aligned and seated on the PCB, solder the remaining pins.



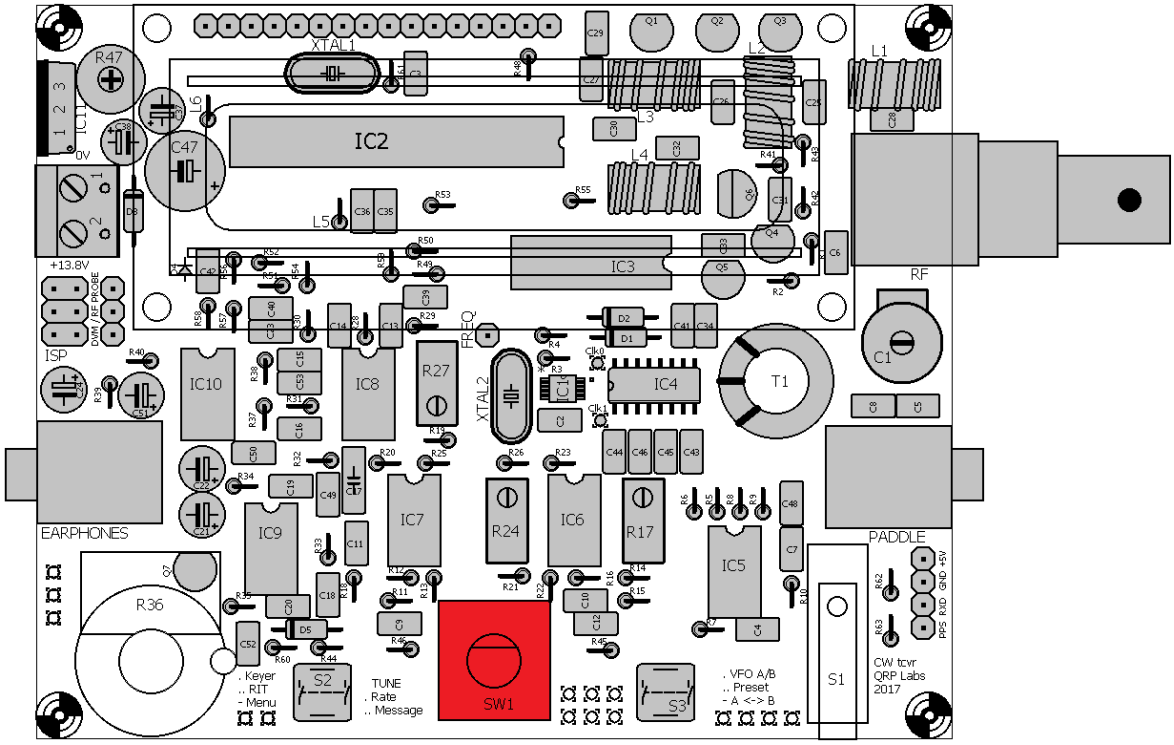
3.58 Install buttons S2 and S3 (left and right buttons)

These buttons (or other push-to-make buttons) may alternatively be wired off-board, for example when mounted in a front panel. Refer to later section for wiring details.



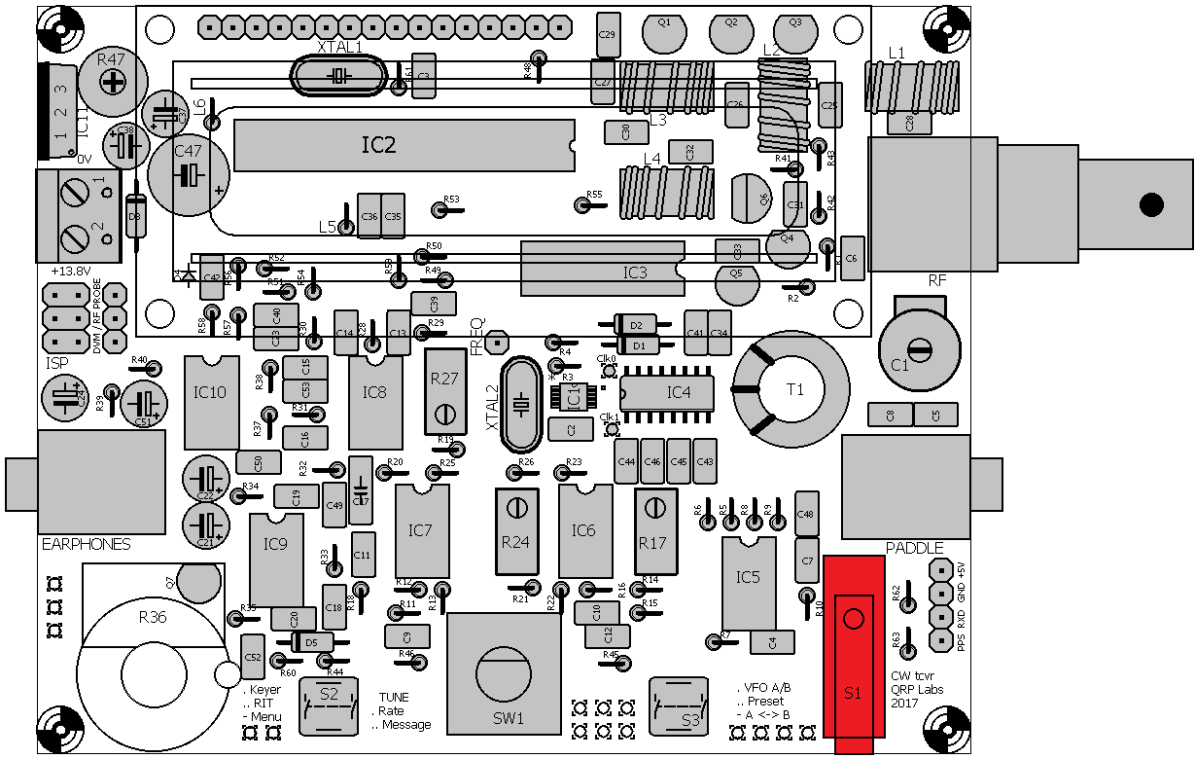
3.59 Install rotary encoder

This switch has seven solder pads: five are for electrical connections and the two large tabs are for mechanical stability. You may also choose to mount the switch off-board, for example on a front panel. Refer to the later section describing the wiring. Remember that it will be difficult to remove the switch later, so you may wish to make the external mounting decision now.



3.60 Install microswitch

The microswitch can be used as an on-board “straight key” if desired and is suprisingly effective! The PCB holes are large so be sure to use enough solder to make good joints.



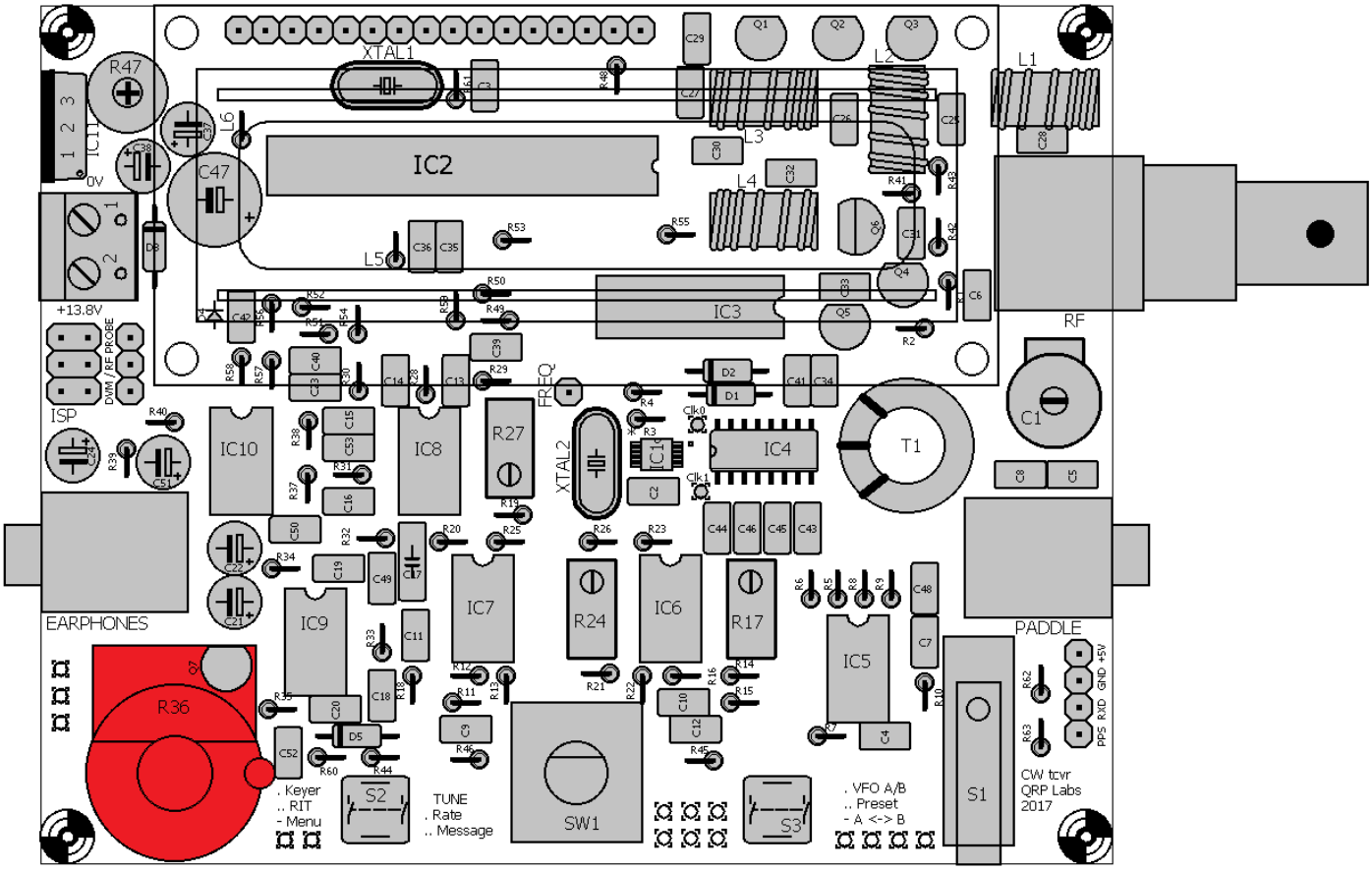
3.61 Install gain control potentiometer R36

Carefully bend the pins of the potentiometer upwards 90-degrees as shown in the photograph.

Remove the nut from the potentiometer, fit the shaft through the hole in the PCB, and carefully install the bolt and tighten with pliers. There is an anti-rotation tab on the metal case, which fits in the smaller hole to the right of the main shaft hole. The tab hole is near to C52, on the right of the potentiometer. When tightening the nut ensure that the pins remain aligned with the corresponding pads on the PCB.

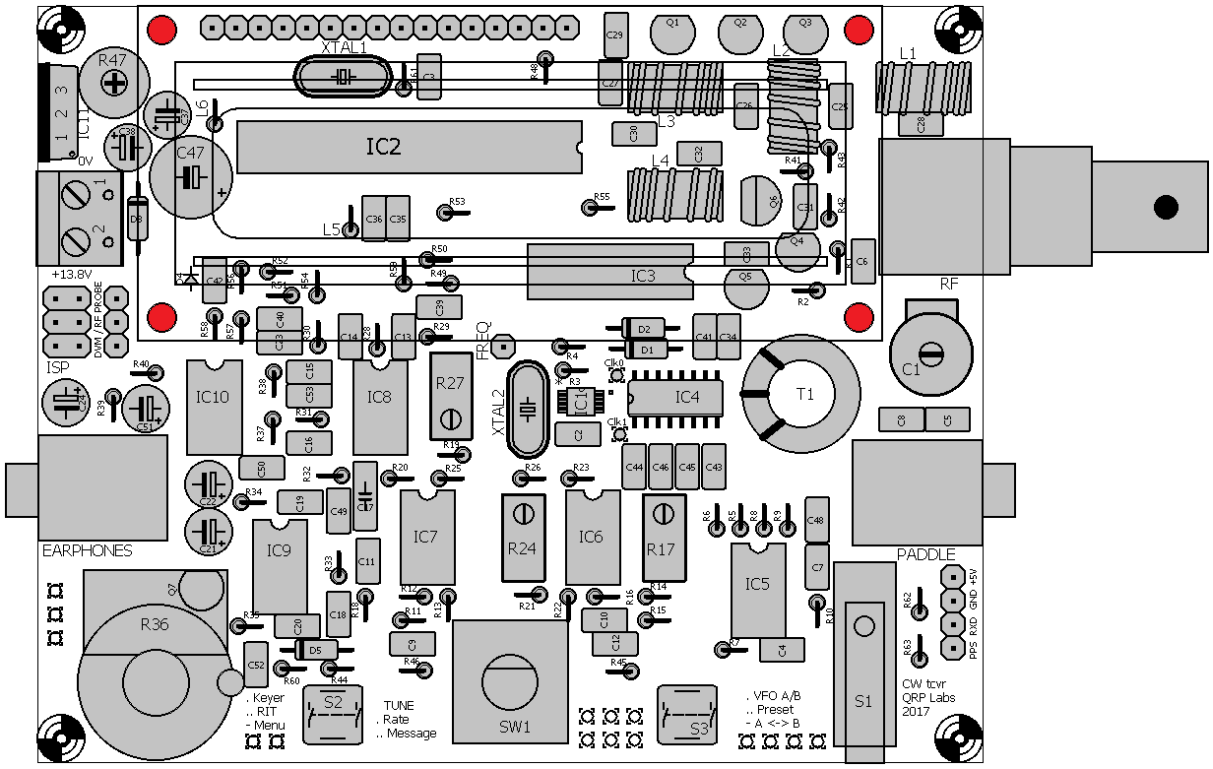
Be very careful with this potentiometer and do not over-tighten the nut. The potentiometer is quite delicate.

The three pins do not extend far enough to fit into the three corresponding holes on the PCB. However they are close enough that a generous blob of solder is enough to make a reliable connection to each pin.



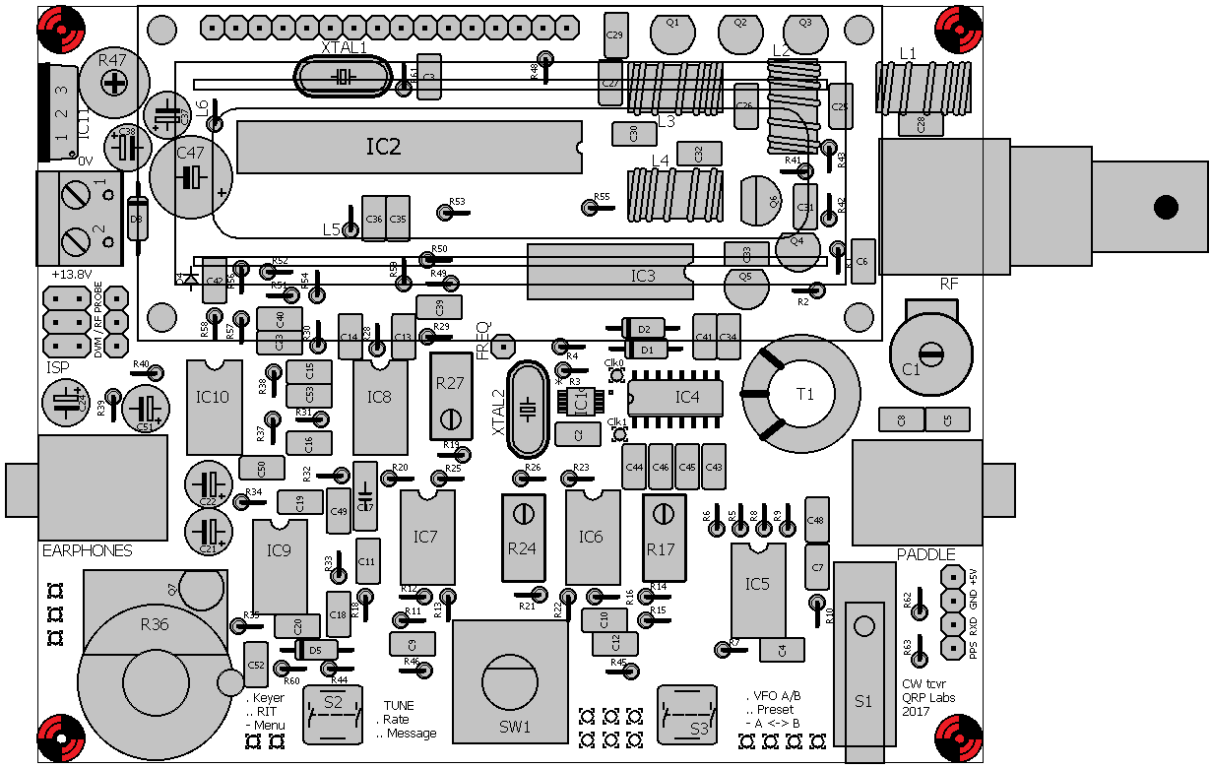
3.62 Install four 12mm hex spacers at the LCD corners

Install the four 12mm spacers on the top side of the PCB, threading four M3 nylon screws through from the underside of the PCB. These four spacers fix the LCD module to the PCB assembly.



3.63 Install four 12mm hex spacers at the PCB corners

These spacers are installed on the bottom side of the PCB, with the four M3 screws threaded through from the top (component) side. These four spacers make four little “pillars” at the board corners, to keep the underside suspended above your workbench or fixing to an enclosure.



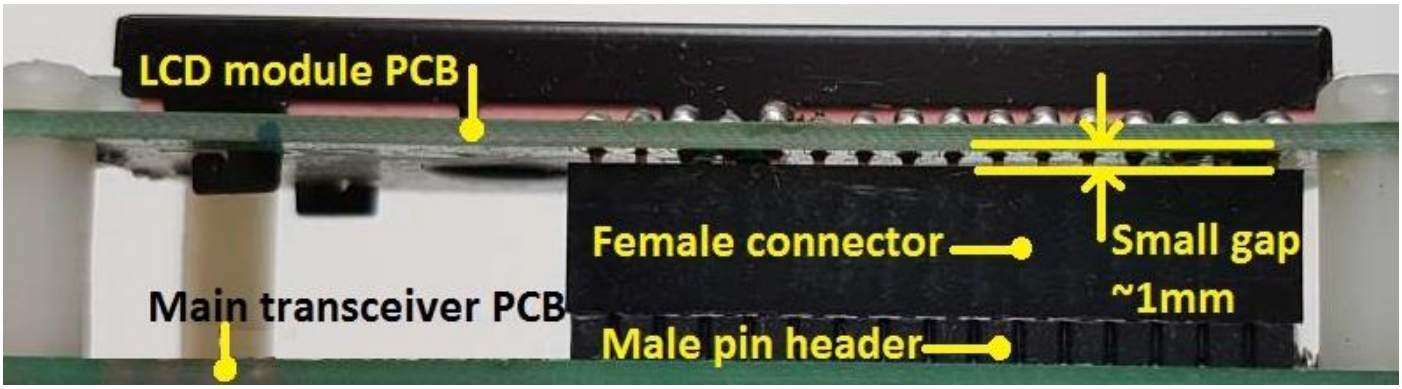
3.64 Install knobs

Two black knobs are supplied, one for the rotary encoder and one for the gain control. When fitting the gain control knob, you should ensure that when the potentiometer is turned all the way anti-clockwise, the white pointer on the knob points to the bottom left corner of the PCB. This is the conventional alignment for volume control knobs.

3.65 Install 16-pin female connector on LCD module

The 16-pin female header connector must be installed on the solder-side of the LCD module, so that it will mate with the 16-pin male header on the main transceiver PCB. The easiest way to do this is to temporarily assemble the modules and bolt the LCD in place using the supplied nylon M3-sized screws.

The nylon headers at the LCD module corners are a little longer than would be ideal. Accordingly leave a small gap between the 16-pin female header connector, and the LCD module itself. Push the female header connector all the way onto the 16-pin header on the main PCB, and solder the pins in place.



3.66 Install microcontroller

Finally install IC2, the programmed ATmega328P microcontroller. Be very careful to ensure that the dimple on the chip is aligned with the dimple on the 28-pin DIP socket, which itself should already be aligned with the dimple on the PCB silkscreen.

3.67 Install LCD and the remaining 4 screws to bolt LCD module

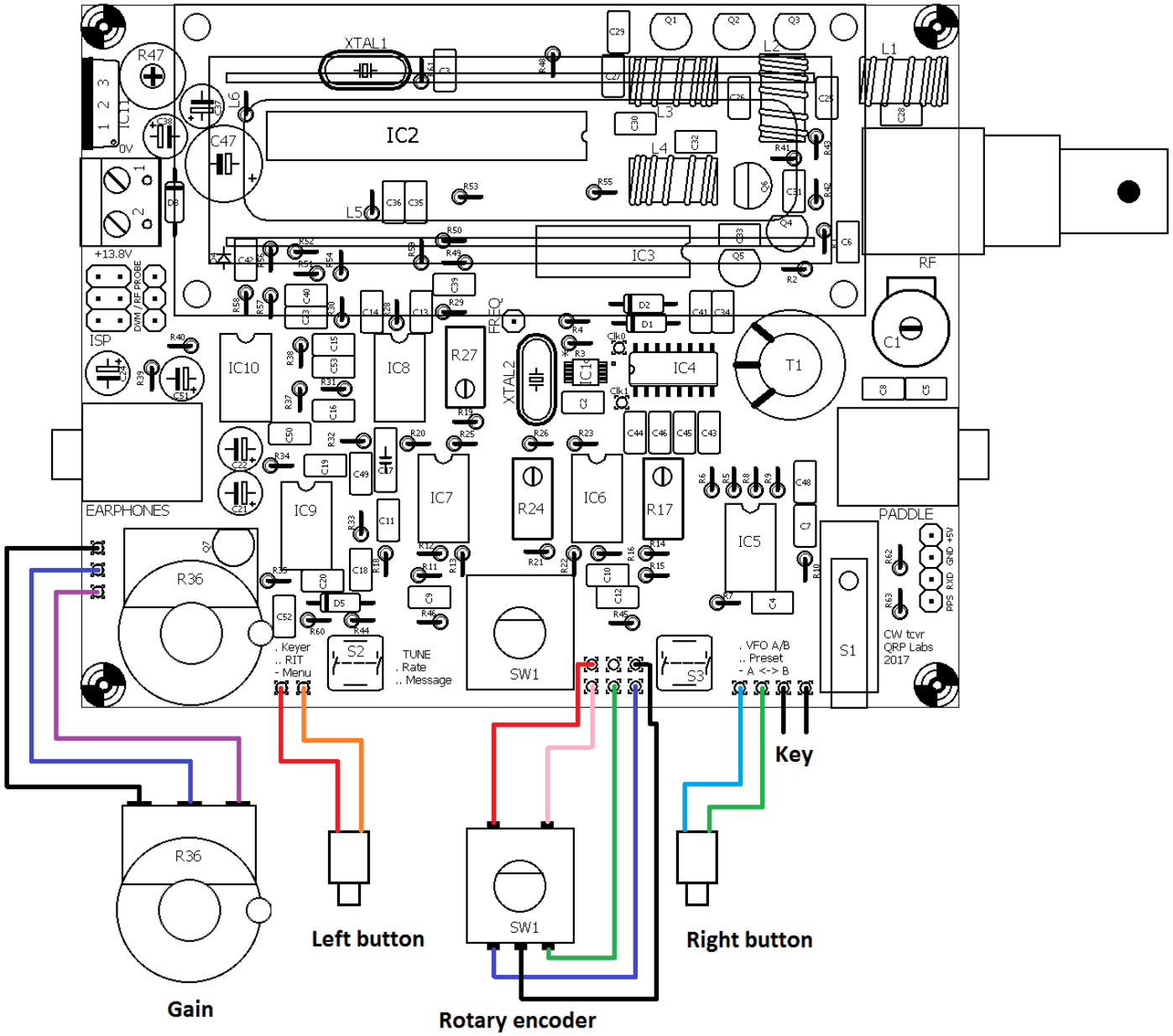
Now plug in the LCD module and screw it in place, using the remaining 4 nylon M3 screws.

3.68 Connections for off-board mounting of controls

The following wiring diagram shows the connections between the PCB pads and controls (buttons, rotary encoder, gain potentiometer) that should be made if you intend to wire the controls off-board. For example, you may wish to install the radio in an enclosure and mount the controls on a front-panel.

The diagram shows the two buttons, gain potentiometer and the rotary encoder. PCB pads are also available for the microswitch “Morse key” as labelled in black (bottom right) but typically you would probably wish to plug a key or paddle in to the 3.5mm socket.

If installing the radio in an enclosure you would also want to wire the LCD module to the PCB rather than plug it in directly. Take care to ensure short cables, or use shielded (screened) cable in order to avoid radiating digital noise into the RF circuits.



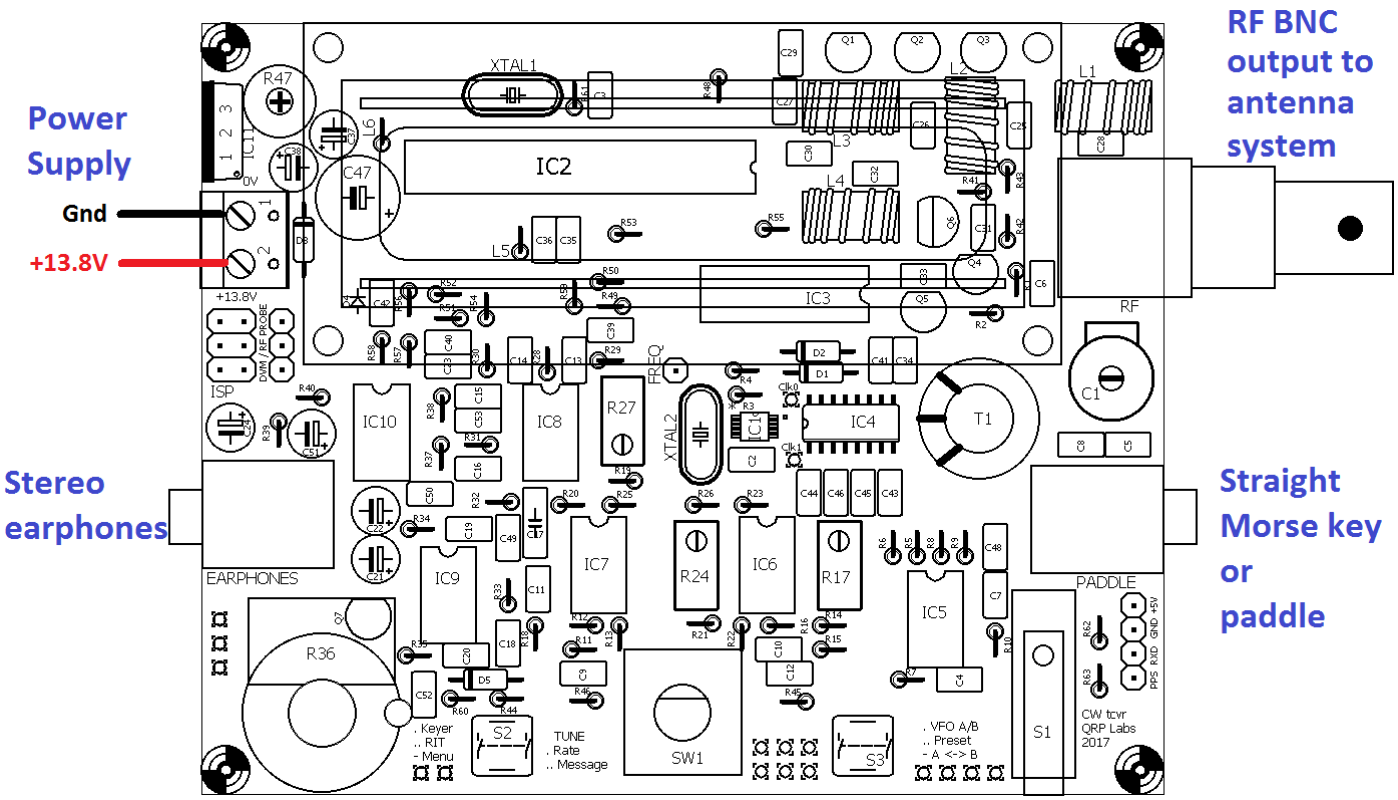
3.69 Connections for basic operation

The following diagram illustrates the basic connections for transceiver operation. A power supply is required, which needs to be able to supply up to 0.5A on transmit. The supply voltage may be from 7 to 16V, and the RF power output will depend on the supply voltage (higher output power is produced at higher supply voltages).

Note that the 7805 voltage regulator requires no heatsink; however the heat production will increase as supply voltage is increased, and the IC does get uncomfortably hot to touch at 16V supply. Although I still do not believe it is necessary, you may feel less disturbed if you fit a small heatsink to the 7805 voltage regulator if you intend to operate at higher supply voltages.

The earphones can be any stereo earphones such as commonly used with audio equipment, mobile phones and so on.

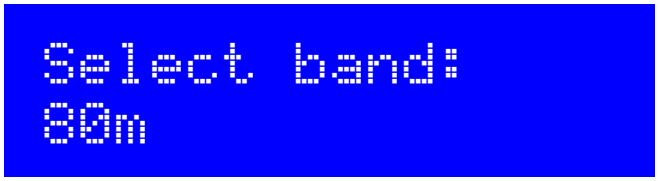
The RF output is a filtered 50-ohm output for connection to a usual antenna system (antenna, and matching unit if applicable).



3.70 Adjustment and alignment

The first thing that you will notice when you apply power to the radio, is that there is probably nothing at all shown on the display. This is because you need to adjust the contrast trimmer potentiometer R47 at the top right of the PCB! Adjust it with a screwdriver until the display text looks right to you.

You should now see the following text on the display:



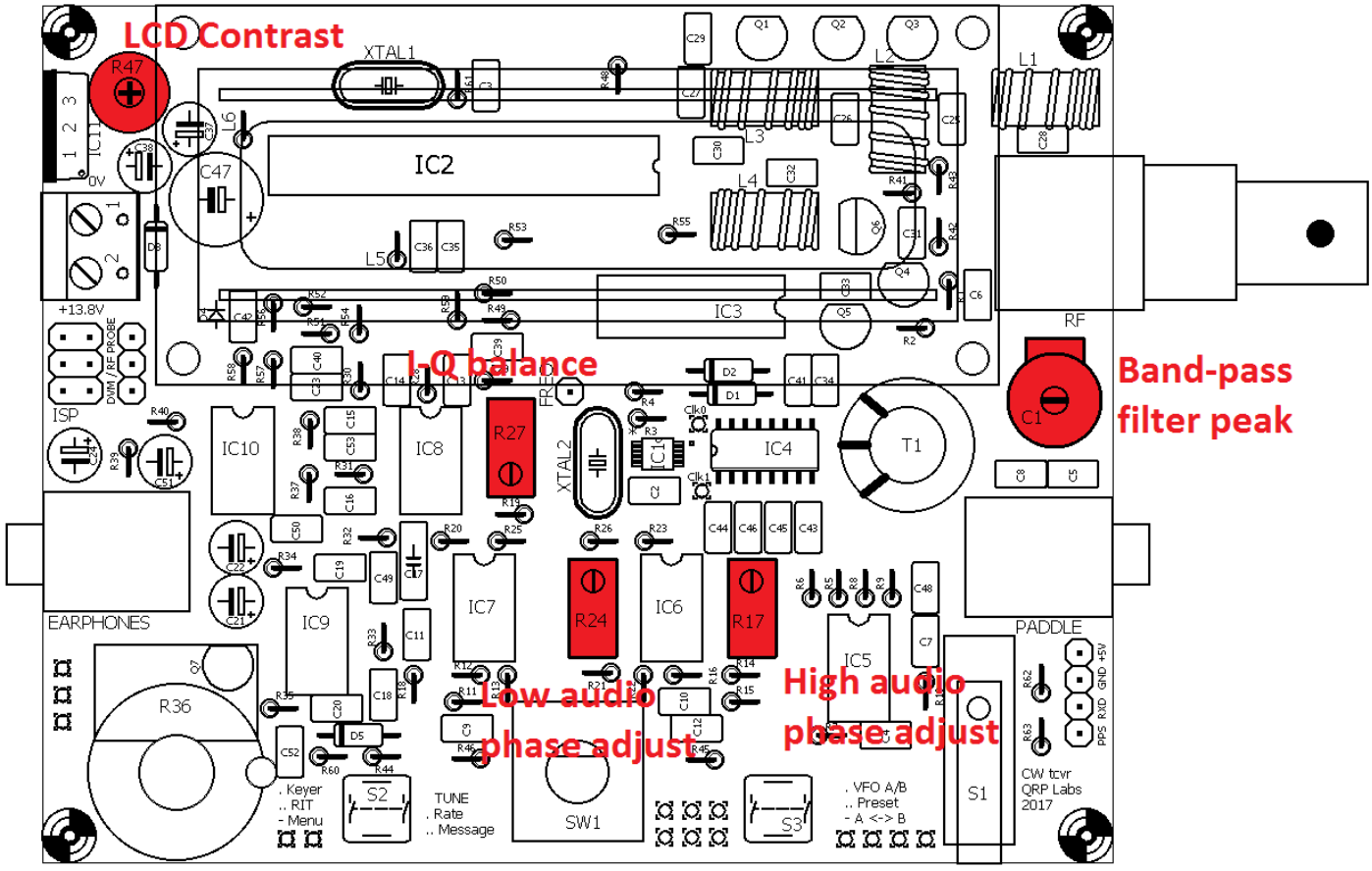
Turn the rotary encoder knob to select the band you have built the kit for. Then press the small button to the left of the rotary encoder, to make your selection.

There are four more adjustments which now need to be made as part of the alignment procedure. The adjustments are:

- Band-pass trimmer capacitor peaking, C1
- I-Q amplitude balance, R27
- Audio phase shift adjustments, R17 and R24

Until these are adjusted, the sensitivity of the radio will be very low. So, do this first, before going any further!

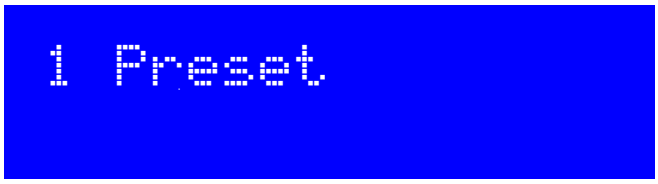
The location of these five adjustments is indicated on the following diagram.



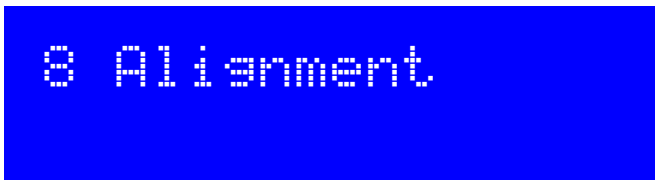
In summary: the alignment tools built into the radio consist of a signal generator which injects a signal into the RF front end, and digital signal processing which adds a 250-Hz digital filter to the existing 200Hz analogue filter, and calculates the amplitude of the signal detected in that bandwidth. During alignment, the amplitude is displayed on-screen as an intuitive bar across the bottom row of the display. Using a screwdriver, you adjust the trimmer component in order to maximise or minimise the displayed amplitude.

Unplug the antenna during alignment of the radio!

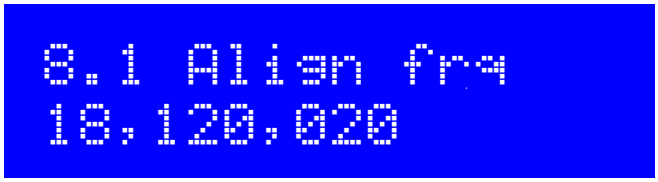
First adjust the band-pass trimmer capacitor C1. To do this, give one long press to the left button. The screen now displays the first menu category:



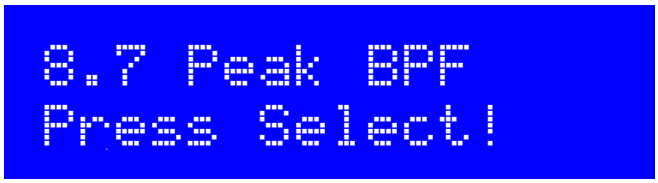
Turn the rotary encoder until you see the alignment menu:



Now press the left (“select”) button, to enter the alignment menu. For example, for 17m operation, the alignment frequency menu item should already be set to a frequency in the CW section of 17m, as follows:



Now turn the rotary encoder until you see:



Do as it says! **But if you have earphones plugged in, please take them out of your ears first.** The tone will be very loud. Press the left (“select”) button to switch on the signal generator so you can adjust the C1 trimmer potentiometer. Now if you have earphones plugged in (and lying hopefully, on your bench), you will hear a loud tone at 700Hz. The display will look like this photograph:

Adjustment of the C1 trimmer capacitor should change the size of the amplitude bar. You need to adjust the C1 trimmer for MAXIMUM amplitude. When this is done, the peak of the band pass filter will be centred on the CW section of the band.

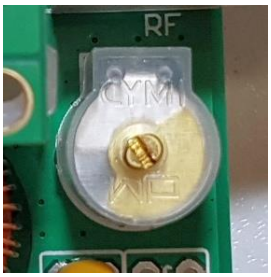
It is very important to understand the number at the top right of the LCD, here shown as 09. This is an amplitude scaling factor, expressed as a power of 2. In this example, the actual amplitude is divided by a factor of 512 (2 to the power of 9) then displayed on the screen. In this photograph, 27 little vertical bars are shown, which means the actual measured amplitude value is 13,824.

If the displayed bar drops below one third of the width of the LCD, then the division factor is reduced by one and the bar is re-displayed. On the other hand, if the displayed bar overflows the right edge of the screen, the division factor is increased by one. This simple method creates an auto-scaling display of the amplitude.

Therefore, to peak the band pass filter trimmer, first adjust the trimmer potentiometer while looking at the division factor in the top right of the LCD. Then carry out the fine adjustment using the displayed amplitude bar. The peak is quite sharp.

Be wary because there can be more than one peak (more than one response of this simple band pass filter). So, tune the trimmer capacitor through its whole range, and determine the maximum scaling factor that you see. In my case here, it is 09. Then make the very fine adjustment necessary to peak the amplitude bar.

When you have peaked the response, check carefully that the C1 trimmer capacitor is not at either end of its range. If it is, this means the resonant circuit is NOT correctly peaked. You need to adjust the number of turns on the long secondary winding of the T1 transformer. Visual inspection will show clearly whether or not the trimmer capacitor is at the end of its range.



Perfect!

OK! The plates of the trimmer capacitor are somewhere nicely in the their range, not at the minimum or maximum capacitance. You have found the peak response of the BPF and all is well.

Plates completely closed

Here the fixed and movable sets of capacitor plates are completely meshed, resulting in the highest capacitance. It means MORE inductance is needed. So carefully unsolder one end of the secondary 3 winding of T1, join a piece of wire, and wind it 5 more turns through the toroid. Then try again. Don't worry if it looks a bit messy.

Plates completely open

Here the fixed and movable sets of capacitor plates are completely unmeshed, resulting in the lowest capacitance. It means LESS inductance is needed. So carefully unsolder one end of the secondary 3 winding of T1, remove 5 more turns through the toroid, and resolder. Then try again.

Adjustment of I-Q balance

Now turn the rotary encoder “one click” clockwise to measure I-Q balance.

Note that the previous alignment used the audio signal BEFORE the final amplification stage, so the gain control had no effect on the signal level. In contrast, the I-Q balance and audio phase shift adjustments use the audio signal AFTER the audio amplification stage. This is necessary because these alignment adjustments inject a signal into the opposite (unwanted) sideband, and the signal level is much lower, therefore it needs to be amplified for the microcontroller to be able to measure it accurately. In this case therefore, the gain control DOES now have an effect. I suggest adjusting the gain control approximately to mid-way to start with. This will provide enough gain, yet not so much gain that the amplifiers are driven into overload, which distorts the signal and measurements.

You should try to carry out the adjustments with the measured audio values in the range 5 to 10 (as indicated by the division ratio in the top right of the display). If it reaches 12, the operational amplifiers are limiting, clipping the signal which will make it difficult or impossible to make the adjustment accurately. If the displayed value (top right of the display) is only 2 or 3, that indicates the gain is too low. Therefore, adjust the volume control so that the display is something like 9.

The I-Q trimmer potentiometer is R27. It is a multi-turn trimmer potentiometer so it may need to be turned quite a few times to get to the optimum value! For this adjustment, you are looking for the MINIMUM amplitude, not the maximum we adjusted the BPF trimmer to. We adjust for minimum because now the injected signal is measuring the unwanted sideband. We want to MINIMISE the unwanted sideband level.

Adjustment of 90-degrees audio phase shift

Similarly turn the rotary encoder one more “click” clockwise, which automatically sets the unwanted sideband audio signal to appear at 600Hz, and adjust the “low audio phase shift” trimmer potentiometer, R24. Again, adjust it for MINIMUM signal.

Turn the rotary encoder clockwise one more “click” and adjust the “high audio phase shift” trimmer potentiometer R17, again for minimum signal.

Now it is necessary to go back and forth between these three menu items for minimum unwanted sideband:

- 8.8 I-Q Bal (adjust R27)
- 8.9 Phase Lo (adjust R24)
- 8.10 Phase Hi (adjust R17)

This is because to some extent, these adjustments influence each other. Obtaining the optimum set of adjustments is an iterative process. So, turn the rotary encoder a click at a time anti-clockwise or clockwise, back and forth through these three menu items. Each time make small further adjustments to the appropriate trimmer potentiometer and observe the lower amplitude. Keep doing this until you see that you cannot really manage to get the unwanted sideband any lower in any of those adjustments. Pressing the right (exit) button twice leaves the menu system and returns the radio to normal operation.

Other items in the alignment menu relate to the calibration of the 27MHz reference oscillator of the synthesiser, and the 20MHz system clock oscillator of the microcontroller. These adjustments can be made manually, or by connection of a GPS module such as the QRP Labs QLG1 GPS receiver kit. However, since this calibration is a lot less urgent than the Band Pass Filter peaking and

unwanted sideband cancellation, they are left until the description of these menu items in the operating manual.

Following the adjustment of these alignment trimmers, the radio is ready to use. A lot of settings are available in the configuration menu, and you should read the operation manual to understand and make use of all the features!

4. Operating instructions

4.1 Overview of features

This 5W CW transceiver kit includes a large number of features defined by the firmware in the ATmega328P microcontroller chip. The following is a brief summary of the features to explore and enjoy. Read the subsequent sections for more detailed information on these features.

Synthesised VFO with rotary encoder tuning

The VFO is an Si5351A Synthesiser chip, configured by the microcontroller. A rotary encoder tunes the VFO, with a variable tuning rate. The radio includes two VFOs, A and B. You can swap from one to the other, copy the contents of the active VFO to the inactive one, or operate Split (Transmit on VFO A, receive on VFO B). There is also a RIT mode offering a receive offset of up to +/- 9,999Hz. The CW offset is also adjustable, and CW-R (sideband swap) mode is supported.

Memory features

There are 16 frequency presets for your favourite operating frequencies. Each frequency preset can be edited in the configuration menu, or loaded/saved into/from the currently operating VFO.

Message mode

The firmware supports storage of 12 messages. Four of these are 100 characters long, the other eight are 50 characters long. A single button click shows the list of messages to send. Message sending can be configured to send just once, or a configurable number of times, or indefinitely repeating. The interval between transmissions is also configurable. The message feature can be useful, for example, for setting up a repeated CQ call with a pause between repeats, during which you can listen for any answers. As soon as you touch the key to begin transmitting, the automatic message mode is cancelled.

CW Keyer

There is an onboard microswitch which can be operated as a “straight” (traditional) Morse key. The firmware also includes an Iambic keyer, for connection of a paddle. The keyer can be configured to operate in Iambic modes A or B, or Ultimatic mode. The keyer speed is variable both from the configuration menu and via a single button press during operation.

With its solid-state, microcontroller operated transmit/receive switch, the radio can operate in full break-in “QSK” mode, or if you prefer, semi-break in.

CW Decoder

A CW decoder operates in the chip. This can be useful on air, particularly for newcomers to CW, though it is never anywhere near as good in QRM or weak signal conditions as a good CW

operator using the wet-ware between his ears. The decoder has a “Practice” mode too, for you to practice your CW sending without actually generating RF. The decoder can also be used to conveniently enter text in the message storage, or for configuration of any of the other menu items. The decoder can also be switched off if desired.

CW or WSPR Beacon mode

A beacon is included too, and this can be configured to operate in CW mode or WSPR mode. Owners of the QRP Labs Ultimate-series weak signal mode transmitter kits will be familiar with the operation of WSPR. A GPS module such as the QRP Labs QLG1 GPS receiver kit can optionally be connected to this CW transceiver kit to provide frequency and time discipline, as well as setting the Maidenhead locator (from latitude and longitude) that is encoded in the WSPR message.

S-meter and Battery voltage

An S-meter and battery voltage indicator can be enabled for display on the LCD. These are both configurable to your needs. The battery voltage indicator would be useful if you intend to operate the radio portable on battery power.

Built-in alignment tools

One of the nicest features of this CW transceiver kit, is the built-in alignment tools menu. The radio can act as its own signal generator, injecting an appropriate signal into the receiver front end, then measuring the audio amplitude after single-sideband demodulation. The menu items allow you to easily peak the Band Pass Filter adjustment, and to adjust the I-Q balance and audio phase shift adjustments for best unwanted sideband cancellation.

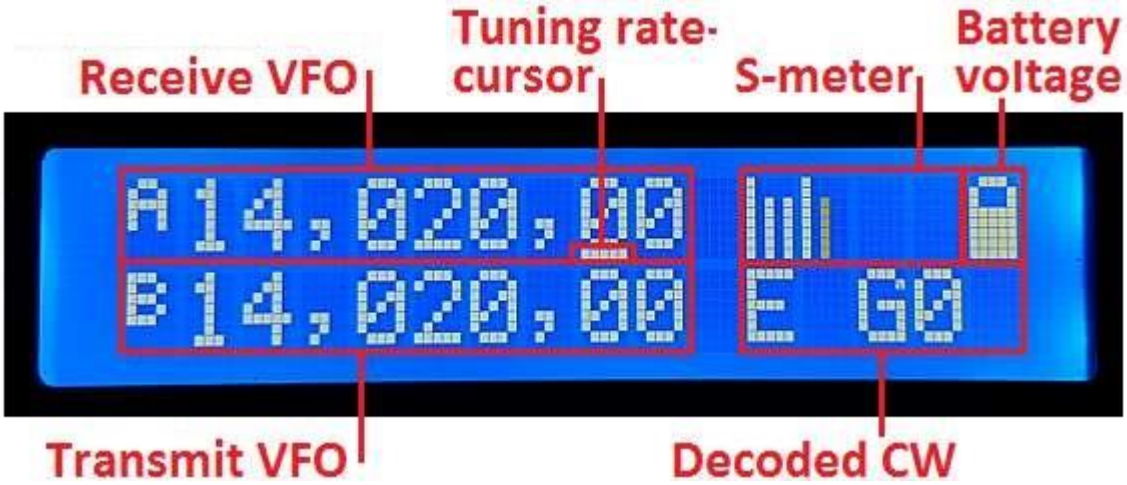
Built-in test equipment

For those times when the assembly didn’t quite go as well as we hoped, and we need to do some debugging – the radio includes its own suite of test equipment too! None of these items of test equipment are going to replace a $100,000 lab any time soon – but they provide very useful measurements for helping people without much test equipment, to debug the radio. They can even be used for testing other projects too!

- Voltage meter
- RF Power meter
- Audio channel amplitude measurements
- Frequency counter (0 to 8MHz)
- Signal generator (3.5kHz to 200MHz)

4.2 Display elements

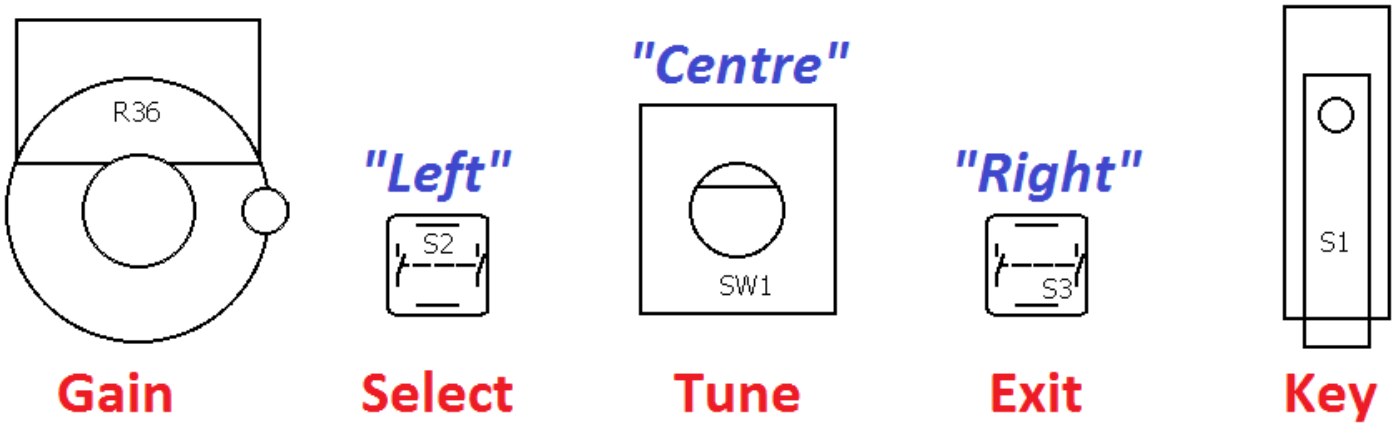
The kit uses a 2 row, 16 character LCD module with a blue backlight.



The main display layout during ordinary operation (which will be called “main operating mode”) is shown in the above photograph. The display during beacon or message transmission modes, menu editing, alignment etc. differs. The main display elements are as follows:

- The receive VFO frequency is always displayed, to 10Hz resolution, at the top left. This may be VFO A or VFO B. The nominally 700Hz CW offset is automatically applied. Ordinarily this frequency is also used for transmission.
- Tuning rate cursor: the underline appears under the digit which is currently tuned by the rotary encoder. In this example, the tuning rate is 100Hz per click, because the cursor is under the 100Hz digit.
- S-meter: these 4 characters display a basic (and not calibrated) S-meter. The scale is configurable (see later description). The S-meter may be shown or hidden.
- Battery voltage: a battery icon appears to indicate the battery voltage in 7 user-definable steps: from full to empty and 5 steps in between. It may also be shown or hidden.
- Transmit VFO: in SPLIT mode, the transmit VFO is displayed on the bottom row of the display.
- RIT (Receiver Incremental Tuning): when not in SPLIT mode, and when the RIT is non-zero, the RIT value is displayed in the bottom left (where the photo shows the VFO B frequency). When RIT is non-zero, and when not in SPLIT mode, the reception frequency is the transmit VFO frequency (which may be VFO A or B) plus the RIT (which may be a negative offset).
- Decoded CW: The remaining space on the bottom row is used for displaying the decoded CW text. When RIT is zero, and you are not operating SPLIT, the whole 16 characters of the bottom row are used for the CW decoder display. However, it may be hidden if desired.

4.3 Operator controls



This diagram shows the operating controls of the radio. All of the controls except the Gain potentiometer have multiple functions, depending on the operating mode, menu editing, etc. The tuning rotary encoder in the centre has a button on its shaft that is activated by pressing it, and this button also has multiple functions.

In this document, the three buttons are mostly referred to as “Left”, “Centre” and “Right”. The original names of the left and right buttons, “Select” and “Exit” respectively, refer to the common actions during configuration menu editing. Select edits a menu item or steps down into a sub-menu; Exit saves an edited menu item or backs up to the parent menu.

It is well worth remembering while you are familiarising yourself with the operation of the radio, that if you get into any unexpected function or menu – you can nearly always press the right (exit) button to cancel and get back to the main operation mode.

Each of the buttons (left, centre, right) may be pressed once, pressed twice (a double-click) or pressed for a long-hold. This facilitates three different functions for each button. In practice, the centre button is stiffer and it is difficult to achieve a double-click; therefore the centre button double-click has no function.

4.4 Tune rate

The rotary encoder tunes the active VFO. The rate of tuning is indicated by the underline cursor. In the example below, the underline cursor is under the comma between the 1kHz and 100Hz digits. This means the tuning rate is 500Hz.

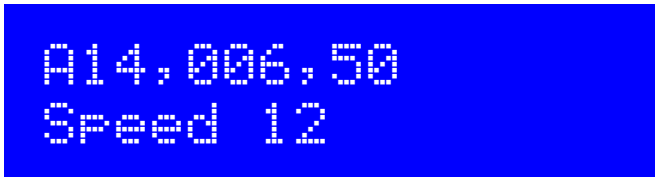
The available VFO tuning rates are 1kHz, 500Hz, 100Hz or 10Hz.

A press of the centre button (in the rotary encoder shaft) causes the tuning rate to change, in the cycle 1kHz -> 500Hz -> 100Hz -> 10Hz -> 1kHz etc.

The default tune rate up at start-up is a configuration item in the VFO menu (see later).

4.5 Keyer speed

The default keyer speed (in words per minute) at start-up is a configuration item in the Keyer menu (see later). During operation of the radio, the keyer speed can be easily adjusted. Click the left button once, and the speed will be displayed on the screen:



Now you can adjust the speed using the rotary encoder. Press any button to return to the main operating mode.

4.6 RIT

RIT (Receiver Incremental Tuning) allows the receive frequency to be adjusted while the transmit frequency (the displayed VFO frequency) remains the same. It is useful if the other station is off-tune, or drifting; other uses include working DX stations who may be listening on a different frequency some kHz away from their transmit frequency.

This radio transceiver allows RIT values from -9,999Hz to +9,999Hz.

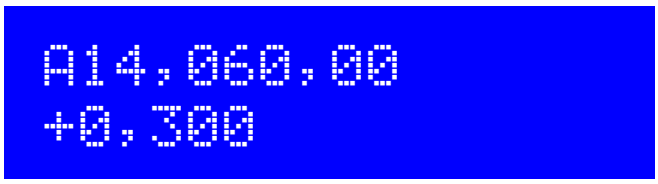
The default RIT at start-up is a configuration item in the VFO menu (see later). RIT can be easily adjusted during ordinary operation by double-clicking the left button:

Now use the rotary encoder to tune the RIT. As you do so, you will hear the RIT immediately applied to the VFO.

The tune rate of the RIT control is again indicated by the underlined digit (here 100Hz). In order to change the tune rate, press and hold the centre button (in the rotary encoder shaft) and turn the rotary encoder at the same time. You will see the cursor move to the left or right 1 digit at a time. Again, the cursor under the comma indicates 500Hz tuning steps.

To cancel RIT tuning (reset the RIT to zero), press the right button (exit); this returns to main operating mode and sets the RIT to zero.

To return to the main operating mode, press the left button (select). Now the RIT is displayed under the VFO, for example:



Remember that cancelling RIT mode is easy, just double-click the left button to show the RIT editing, then press the right button to cancel it (which means, set it to zero).

4.7 VFO mode

A single press on the right button changes the active VFO mode. There are two independent VFOs named A and B. There are three VFO modes for using these VFOs:

- VFO A is active as transmit and receive VFO; if non-zero, RIT is applied during receive
- VFO B is active as transmit and receive VFO; if non-zero, RIT is applied during receive
- Split: VFO A is used for receive, VFO B is used for transmit; RIT is ignored completely

Split mode is often used by DX stations, they transmit and receive on separate frequencies.

4.8 VFO A/B frequency swap

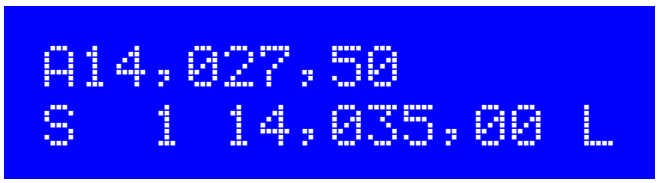
The contents (frequency) of VFO A and B can be swapped by a single long key-press to the right button. This can be useful when setting up the VFO frequencies.

4.9 Frequency Presets

There are 16 frequency presets which may be used for storing your favourite frequencies, or for just temporary use, or however you wish!

The presets are labelled 1 to 16, and can be individually edited in the Preset menu (see later). Often it is more convenient to just save them from the current VFO frequency.

To show the list of preset frequencies in normal operation mode, double click the right button. The display now shows something like this:



The top row of the display shows the currently active VFO frequency as usual.

The bottom row shows a “1” in the 4th character, this is the number of the preset displayed. The next number (here 14,035,00) is the frequency stored in Preset 1.

Use the rotary encoder to scroll through the list of presets until you find the one you want. Once you have selected the desired preset, press one of the three buttons to Save, Cancel or Load the preset, as follows:

- SAVE the current VFO to the selected preset, by pressing the left button
- CANCEL the preset operation (back to main operating mode), by pressing the centre button
- LOAD the selected preset frequency into the current VFO, by pressing the right button

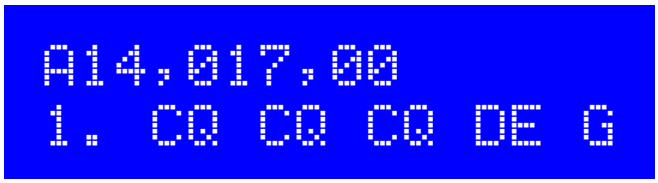
The “S” in the first character and “L” in the 16th character at the far right, are intended as a reminder of which button to press to Save and to Load.

4.10 Automated message transmission mode

My favourite use of the automated message transmission mode is to send a CQ call repeatedly. If a station answers, you can start transmitting and it cancels the automatic CQ mode.

There are 12 message memories. The first four memories are 100 characters long; the remaining eight memories are 50 characters long.

In order to send a pre-saved message, press the centre button with a single long press. The first of the saved messages is shown on the screen, for example if a CQ call is stored in Message 1, you may have something like this:



The bottom row shows the message number at the far left (here it is message 1) followed by the first part of the stored message. If it is blank, that means of course that you have not stored any messages yet!

You can now use the rotary encoder to scroll back and forth between the 12 stored messages and find the one which you want to transmit.

The message can be transmitted multiple times according to the “Repeats” parameter in the Messages menu (see later description). The interval between the repeated transmissions is also defined in the Messages menu, in the “Interval” parameter.

Once you have selected the message that you want to transmit, press one of the three buttons as follows:

- REPEAT: Transmit the message repeatedly by pressing the left button
- ONCE: Transmit the message only once by pressing the centre button
- CANCEL: Cancel the messages operation by pressing the right button

When REPEAT message transmission mode is activated, the number of repeats and the interval between repeats is as specified by the Repeats and Interval parameters in the Messages menu.

The stored message transmission is sent at the currently defined keyer speed.

During the actual stored message transmission, you can immediately cancel the transmission at any time by pressing the right button (exit) or by keying the transmitter with the onboard microswitch Morse key, or a paddle if you are using one.

4.11 Menu System

There is an extensive menu system with more than 80 different configuration or operating parameters stored in non-volatile memory (EEPROM). These are editable to control every aspect of the radio’s behaviour.

The menus are organised into 9 groups as follows:

- 1. Preset
- 2. Messages
- 3. VFO
- 4. Keyer
- 5. Decoder
- 6. Beacon
- 7. Other
- 8. Alignment
- 9. Test equipment
- Save settings!

To enter the menu system, give a single long press to the left (select) button. Use the rotary encoder to scroll back and forth between the nine sub menu groups listed. To enter one of them, press the left (Select) button. To return to the main operating mode, press the right (Exit) button.

The golden rule while in the menu system, is to press the left (Select) button to go in to a deeper menu level or edit an item, and the right (Exit) button to back up.

In order to edit a menu item, navigate to the menu item then press the left (Select) button to start editing. When you have finished editing the item, press the right (Exit) button to save it.

NOTE that changes to configuration parameters only take effect on the radio, when you leave the menu system and return to the main operating mode. During viewing or editing of menu items, the radio remains in receive mode on the currently selected VFO frequency. The exception to this is certain alignment and test equipment tools, where these need to use parts of the radio.

4.12 Saving current operating parameters (VFO frequency etc)

The final (10th) item “Save settings!” is a special item which causes certain operating parameters to be stored to EEPROM. These then are the defaults the next time you power up the radio. This is very convenient if you want to recommence operating next time you switch on, with exactly the configuration your radio is in now. The controls sequence to save the settings is very simple. This makes it easy and quick to save your current configuration when you wish to switch off the radio.

a) Single long press to the left button, to enter the configuration menu system
b) Turn the rotary encoder one click anticlockwise to show “Save settings!” on the screen
c) Press the left button once more, to actually save the settings

The list of items saved is:

- VFO Mode (A, B, Split)
- VFO A frequency
- VFO B frequency
- Tune rate
- RIT
- RIT tune rate
- Signal generator frequency

4.13 Types of configuration menu item

There are five types of menu configuration item, and editing these is a little different depending on the type.

1) LIST: a fixed list of values applicable to that menu item, for example Keyer mode
2) BOOLEAN: an ON/OFF parameter, such as used to control whether the battery icon shows
3) NUMBER: a numeric parameter such as a stored frequency preset
4) TEXT: a text configuration item such as a stored message
5) UNEDITABLE: some menu items are display-only, such as the alignment or test tools

4.14 Editing a configuration menu parameter

To start editing a parameter, navigate to the desired parameter in the appropriate menu, and then press the left button (Select). When editing is active, you will see a cursor appear under the digit being edited. For example, here is menu item 2.13, the message repeat interval:

The underline cursor below the 4 indicates that editing is active; turning the rotary encoder will change the parameter value.

When you are finished editing, press the right (exit) button to conclude editing. This saves the parameter in the microcontroller's EEPROM memory.

A flashing inverted cursor is also available, you can choose that in the "Cursor Style" parameter in the "Other" menu (see later section).

4.15 Editing a LIST parameter

Editing a list parameter is very simple, it is just a matter of turning the rotary encoder. The display scrolls through the list items. For example, this is the Keyer mode parameter, menu 4.1:

Note that the editing indicator cursor appears under the leftmost character.

When you are happy with your selection, press either the left or right buttons, to save the change.

4.16 Editing a BOOLEAN parameter

Editing a BOOLEAN parameter is exactly the same as editing a LIST parameter, except that now the list of items is always restricted to just ON and OFF (representing True/False).

4.17 Editing a NUMBER parameter

When editing a number parameter, the cursor underline appears under the currently edited digit. The cursor starts at the far left (most significant digit). The rotary encoder tunes the digit. The operation is very similar to tuning a VFO in ordinary operation. This example shows editing the default (power-up) VFO A frequency:

To alter the “tuning rate”, you can either

a) Press the left (select) button to move the cursor to the next digit to the right OR
b) Press and hold the centre button, while turning the rotary encoder shaft – this allows you to move the cursor left or right.

Editing of the number is concluded, and the number is stored to EEPROM, when either:

a) You press the right (exit) button OR
b) You press the left (select) button so many times that the cursor falls off the right hand side

Yet another way to input numbers, which is a REALLY convenient way, is to use the Morse key and CW decoder! During editing of numeric parameters, the CW decoder is activated but only decodes number characters 0-9. The CW decoder expects well-timed CW, with correct spacing between words and characters. The CW decoder expects you to key in the numbers at near to the configured Keyer speed. If you start keying at a much different speed, the CW decoder will adapt to your keying but this may take several characters to “sense” your keying speed, so some characters can be missed.

When you have keyed in the whole number, the number is automatically saved to EEPROM, which leaves editing mode. Once you are used to editing numeric parameters by keying in CW, it becomes the easiest and fastest way to edit menu parameters.

4.18 Editing a TEXT parameter

An example of a text parameter you may wish to edit is the stored messages. For example, stored message 2 is edited in menu item 2.2:

By far the easiest way to edit TEXT parameters is simply to use the CW decoder! As before, it expects well-timed CW, with correct spacing between words and characters. The CW decoder expects you to key in the characters at near to the configured Keyer speed. If you start keying at a much different speed, the CW decoder will adapt to your keying but this may take several characters to “sense” your keying speed, so some characters can be missed.

Editing of the parameter is concluded either when you press the right button (exit), or when no more characters are available for editing; for example, if you filled up the chosen Message memory.

It is also possible to edit a text parameter entirely with the buttons and rotary encoder, though this is usually a slower way to edit text parameters. Owners of the QRP Labs Ultimate3S (or earlier) QRSS/WSPR transmitter kits will already be familiar with this style of editing text.

The text parameter supports all of the characters which the Message keyer can encode, which is the same as the CW decoder can decode. Specifically, A to Z, 0 to 9, Space, then punctuation characters / \ ? . , (note, the break character \ appears at the moment as the Japanese yen character ¥ - this will be fixed in a future firmware version).

The following characters/symbols have special functions.

Insert: Use this symbol to insert a character in the text. Find this character using the rotary encoder, then press the left button to activate it. All the characters to the right of the cursor position are shifted right one position, including the character which was originally in the current position.

Backspace (delete): If you select this character as the current character using the rotary encoder, then when you press the left button, the current character is deleted and the cursor moves back left one position.

Delete all: If selected as the current character, pressing the left button has the effect of deleting the entire message, starting again at the left of the screen. There is no “undo”, so use with caution!

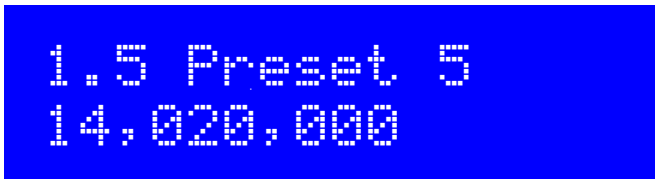
Enter Right (finished): The behaviour of this symbol is the same as Enter, except that it preserves all the text, including the text to the right of the cursor. It simply saves the whole line.

Enter (finished): If selected as the current character using the rotary encoder, pressing the left button is used to finish editing the setting. The setting is saved, and you leave editing mode. Note that the text that is saved is only the text to the left of the Enter symbol. If you select this symbol and press the left button when you are not at the furthest right position of the message, then everything right of your position is deleted.

You can also move the cursor backwards and forwards within the text being edited, by pressing and holding the centre button, then turning the rotary encoder. This moves the cursor position within the text parameter.

4.19 Frequency presets menu

There are 16 frequency presets, labelled 1 to 16. This example shows Preset 5:



All of the Preset menu items are NUMBER types. Refer to the “Editing a NUMBER parameter” section above for instructions on how to edit a NUMBER parameter.

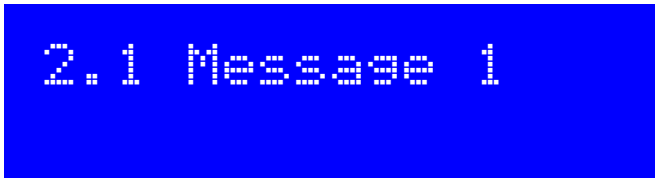
It is also convenient to load the current VFO into the preset memories as described in the section above titled “Frequency Presets”.

4.20 Messages menu

There are 14 configuration items in the Messages menu:

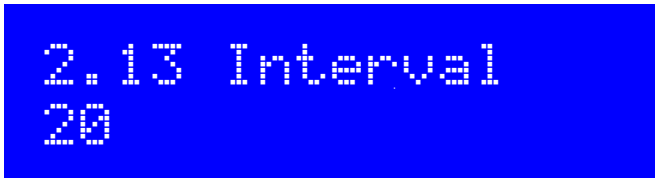
2.1 Message 1 (100 characters text stored message)
2.2 Message 2 (100 characters text stored message)
2.3 Message 3 (100 characters text stored message)
2.4 Message 4 (100 characters text stored message)
2.5 Message 5 (50 characters text stored message)
2.6 Message 6 (50 characters text stored message)
2.7 Message 7 (50 characters text stored message)
2.8 Message 8 (50 characters text stored message)
2.9 Message 9 (50 characters text stored message)
2.10 Message 10 (50 characters text stored message)
2.11 Message 11 (50 characters text stored message)
2.12 Message 12 (50 characters text stored message)
2.13 Interval
2.14 Repeats

The 2 stored message presets are 100 or 50 characters long, according to the list above. For example:

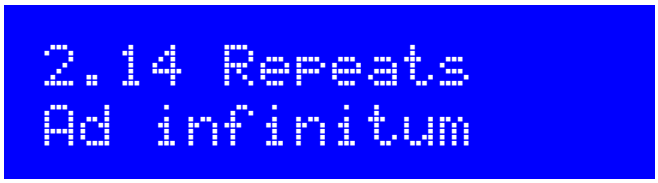


Turn the rotary encoder to select the Message from 1 to 12 which you wish to edit, then press the left (“select”) button. Now you can edit the message text in one of two ways:

1) Choose each character individually from the list, using the rotary encoder. When you have chosen the correct character, press the left (“select”) button to move to the next character. This process is described in more detail in the preceding section on editing text parameters.
2) Key in the desired text on the straight Morse key on the board, or using your external paddle. The CW decoder must be enabled for menu editing (see “Enable edit” parameter).



The Interval is a NUMBER parameter that specifies the interval in seconds, between repeated transmission of a stored message (if repeats are configured: see next parameter).

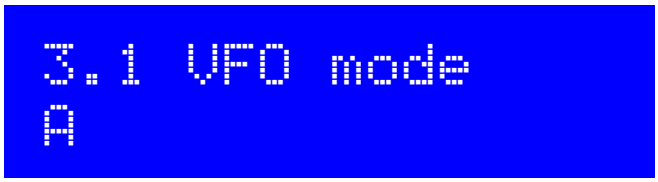


The Repeats parameter specifies how many times the message transmission will be repeated, in the repeat transmission mode. The Repeat parameter is actually a LIST type, since it can be a number from 1 to 99, or “Ad infinitum”. In the latter case, the Message transmission continues indefinitely (ad infinitum).

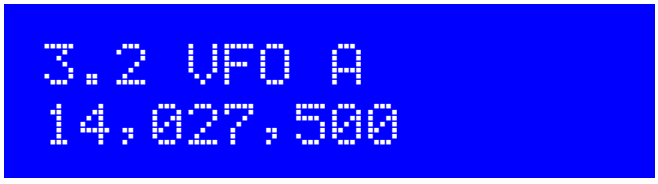
4.21 VFO menu

The VFO menu contains a number of configuration parameters which affect the VFO operation and CW reception. Many of these parameters are only power-up default parameters for the VFOs. The current VFO frequency, mode etc is not stored in these parameters. If you wish to save the current VFO frequency, mode etc into EEPROM so that when the radio is next powered up, the exact same state is available from power up, use the “Save settings!” feature described previously.

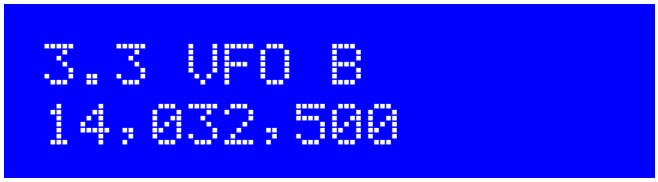
These configuration items in this menu are explained below.



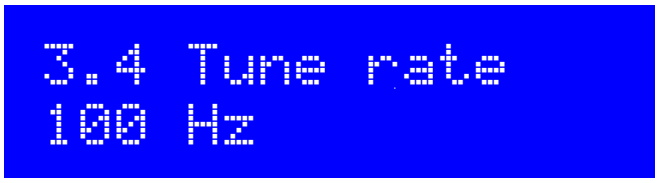
This parameter specifies the VFO mode at power-up. It can be A, B or Split. The current VFO mode during ordinary operation is not stored in this parameter unless you “Save settings” as described previously.



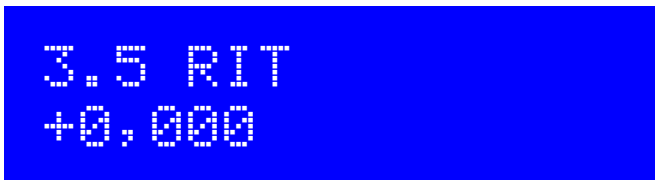
This parameter specifies the VFO A frequency at power-up. The current VFO A frequency during ordinary operation is not stored in this parameter unless you “Save settings” as described previously.



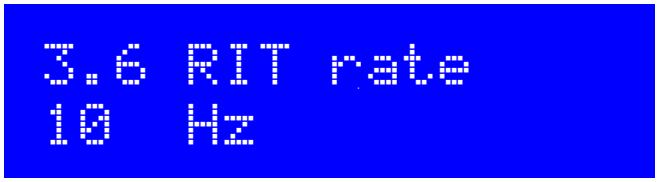
This parameter specifies the VFO B frequency at power-up. The current VFO B frequency during ordinary operation is not stored in this parameter unless you “Save settings” as described previously.



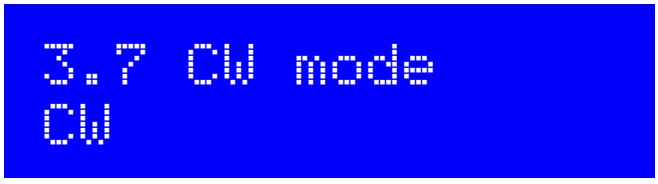
This LIST parameter specifies the Tune rate power-up. Available values are: 10MHz, 1MHz, 100kHz, 10kHz, 1kHz, 500Hz, 100Hz, 10Hz or 1Hz. The current Tune rate during ordinary operation is not stored in this parameter unless you “Save settings” as described previously.



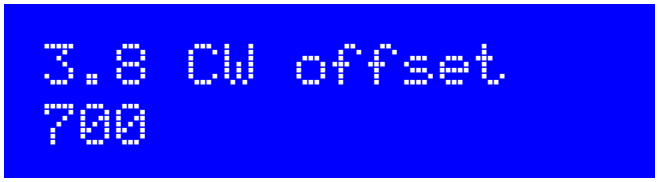
This parameter specifies the RIT at power-up. The current RIT amount during ordinary operation is not stored in this parameter unless you “Save settings” as described previously.



This parameter specifies the RIT tuning rate at power-up. The current RIT tuning rate during ordinary operation is not stored in this parameter unless you “Save settings” as described previously.



This LIST parameter specifies the CW mode. Ordinarily CW is received in Upper sideband with a 700Hz offset. They may be some occasions where operation on the other sideband is desired (lower sideband), for example to exclude an interfering nearby station when the CW filter performance is asymmetric. In these cases you can change the parameter value to CW-R (reverse mode CW) to select Lower sideband reception mode.



This parameter specifies the CW offset in Hz. It is the difference between the VFO frequency during transmit and receive. It is automatically applied during receive in order to resolve a perfectly netted transmission on the same frequency as yours, at 700Hz audio output.

If you wish to change the CW offset frequency you may do so using this parameter. However, be warned that the 200Hz audio analogue filter in the circuit is still centred on (approximately) 700Hz. If you moved the configured CW offset frequency too far away, then the received audio will fall outside the CW filter and be attenuated.

The chosen band configuration for your radio. On first powering up the radio, the display requests your selection of band, and it is stored in this configuration parameter. You are not likely to need to touch this setting again. Changing the band results setting sets a frequency in the chosen band, 20kHz above the lower band edge. This frequency is written into:

- VFO A
- VFO B
- Presets 1 to 16
- Alignment frequency
- Beacon frequency

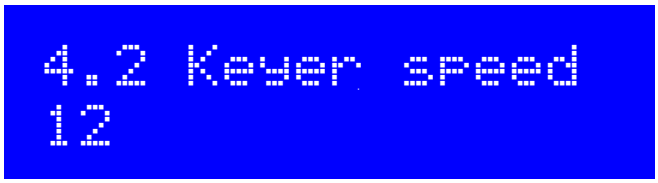
4.22 Keyer menu

The Keyer menu contains a number of configuration parameters relating to the CW keyer, which are described below.

The mode of the CW keyer function in the firmware. Possible modes are:

- Straight
- IAMBIC A
- IAMBIC B
- Ultimatic

If you wish to use a traditional up/down Morse key, or the onboard microswitch, these are called “straight” keys and you should select the “Straight” mode. If you wish to use a modern paddle plugged into the 3.5mm socket on the right side of the transceiver PCB, then select the desired operating mode e.g. IAMBIC A.

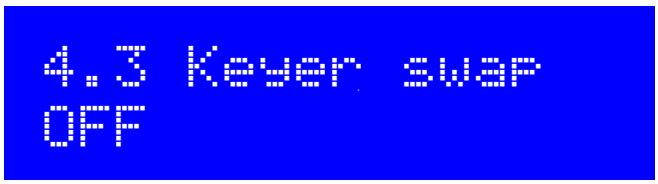


This is the default power-up speed of the Keyer function in Words Per Minute (wpm). The current Speed during ordinary operation is not stored in this parameter unless you “Save settings” as described previously.

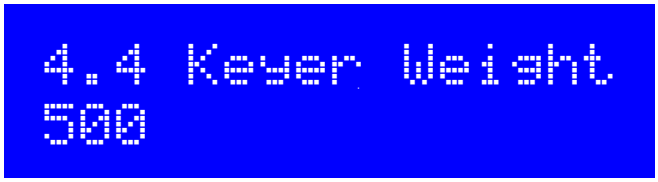
Keyer speed can be altered easily from the ordinary main operating mode, with a single press of the left button, as described previously.

Keyer speed is also used for stored message sending, and for the beacon function when in CW mode.

The keyer speed is also used to configure the CW decoder during transmit or during menu item editing. When transmit or editing are commenced, the speed is copied into the CW decoder and initialises it. Thereafter if you transmit (or enter data during editing) at a different speed, the CW decoder will adapt to your speed. However, if the speed difference is large, the adaption can take several characters to correctly “sense” your keying speed, which can result in incorrectly decoded characters or missed characters.



This is a BOOLEAN parameter which lets you swap the “dit” and “dah” connections in software, if you find that your paddle is reversed.



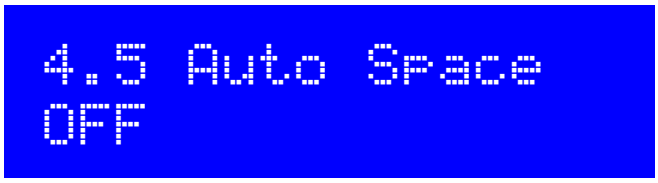
Ordinarily Morse dit and dah durations have a 1:3 ratio. The space between symbols is equivalent to 1 dit, between characters 3 dit lengths, and between words 7 dit lengths. This is standard Morse timing. However, some people may wish to alter this, for various reasons.

The Keyer Weight parameter allows variation of the ratios. The value has three digits. The default value of 500 corresponds to 50.0%. This means the “duty cycle” of a stream of dits is exactly 50%. The key-down dit length is therefore the same as the key-up inter-symbol pause.

If the weight is increased from the default 50.0%, then the key-down “dit” is made longer. A “dah” is lengthened by the SAME amount. The corresponding inter-symbol (or character, or word) gap is shortened by the same amount. The additional time spent on the key-down is therefore taken from the key-up period. The keyer speed is unchanged by altering the weight parameter.

As an example: suppose you want to make your Morse sound “harder” by shortening the dits and dahs. You could set the parameter to 450, which means 45.0%.

The parameter range is 050 to 950 (5% to 95%) though you would not normally ever need to go anywhere near these unreasonable limits. In the event that you enter a value outside this range, the firmware automatically applies these limits to the actually used parameter.



Auto-spacing means that the pause between CW characters is forced to be 3 dit lengths (more or less, if you have CW weighting configured, see above).

The majority of keyers do not implement auto-spacing. You use the paddle to send your dits and dahs making up the Morse character you wish to send. As soon as you next press the paddle, the next character is started. The keyer forces correct 1:3 ratio of dits and dahs and inter-symbol spacing, but it does not force you to wait for the correct duration of 3 dits between transmitted characters.

Some keyers do implement automatic character spacing, such as the old (1973) Accu-Keyer design by James WB4VVF see https://inza.files.wordpress.com/2011/01/accu-keyer.pdf .

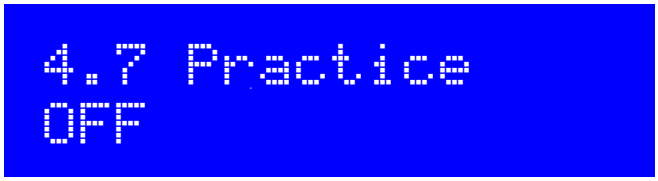
This configuration therefore allows you to switch on automatic character spacing if you wish. In this case, if you press a paddle too SOON, before the 3 dit durations have elapsed after the last character completed, the keyer will wait until the correct time to start the next character.

In the even that you press the paddle too LATE, there is nothing the keyer can do to travel back in time and force it to 3 dit lengths for you. You might have intended an inter-word space, for example. So pressing the paddle too late cannot be corrected.

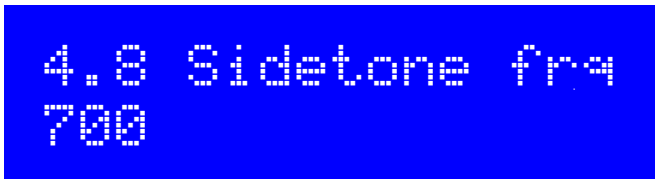
This setting defines the break-in (QSK) behaviour of the radio. Two settings are possible:

Full: after the delay time for RF envelope shaping, the Transmit/Receive switch is set to “Receive” shortly after key-up. In this way, you will hear the other station (or any QRM, QRN etc) transmitting in between the dits and dahs of your own transmissions. Many experienced operators like to be able to have a feel for what is happening on the band, in between their key-downs. In some ways you feel like you are listening to your own sidetone audio as just another signal on the band, and you can still hear other signals too.

Semi: After key-up, there is a delay before the Transmit/Receive switch is set back to “Receive” mode. The receiver is therefore kept muted during your whole CW transmission, not listening to the band in between your transmitted symbols. Many operators prefer to avoid the distraction of hearing the band between their dits and dahs. In Semi-QSK mode the Transmit/Receive switch is set back to “Receive” only after a suitable delay (of 8 dit lengths), long enough to occur only at the end of the transmission.

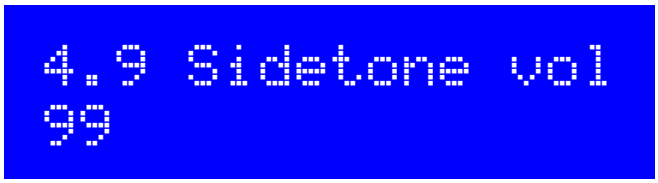


Normally you would leave Practice mode switched OFF. However if you want to practice sending CW, and see if the CW Decoder can decode you, then you can switch Practice mode to ON. In practice mode, the radio does everything it normally would, except that it never sends any RF power to the antenna!



This NUMBER parameter allows you to change the Sidetone frequency if you wish. Sidetone is the audio tone which is generated by the microcontroller on key-down and injected into the audio signal path. Sidetone is ONLY an operator convenience to let you hear your keyed signal, and has no impact at all on the transmitted RF amplitude or frequency.

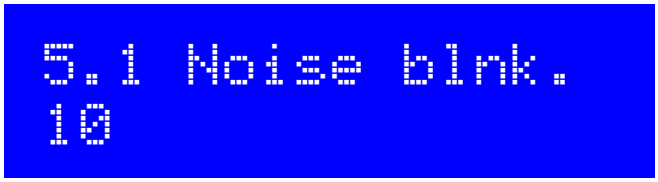
It is strongly recommended to leave the Sidetone frequency set to the same frequency as the CW Offset frequency in the VFO menu. Your ears will get used to recognising the 700Hz audio tone, and it makes it then much easier to accurately tune to a station you hear transmitting on air, such that his signal will also be at 700Hz audio in your receiver. When you hear him at 700Hz, he will be in the middle (near, anyway) of your CW audio filter. When you transmit you will also be netted accurately to his frequency. All being well, this should mean that your signal also ends up nicely in his CW filter passband too. Then you can QSO!



You can use this parameter to reduce the volume of the sidetone audio. Most people will find the default setting of 99 too loud. So experiment with lower values until you find a sidetone audio level which you find comfortable.

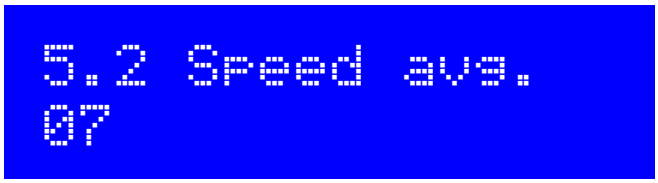
4.23 Decoder menu

The Decoder menu contains a number of configuration parameters relating to the CW decoder, which are described below. Some of these parameters control some aspects of the decoder behaviour. Some constructors may find it interesting to experiment with these settings and see if you can improve the performance of the CW decoder in your specific circumstances. For example, some stations may experience more noise interference than others, depending on your location etc.



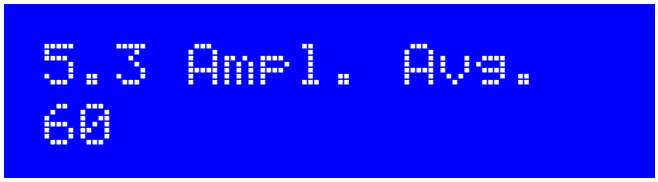
This parameter defines the duration of the noise blanker in milliseconds. The microcontroller's 10-bit ADC samples audio at 12.019 samples per second. 48 samples are analysed by an implementation of the Goertzel algorithm (kind of a single bucket of a Fourier Transform), which results in a digital filter bandwidth of 250Hz. In other words, it results in a measurement of the amplitude 250 times per second, i.e. once every 4 milliseconds. The amplitude is analysed by logic which compares it to a threshold amplitude to decide if a tone has been detected or not. Impulse noise that generates shorter pulses than the noise blanker parameter, is ignored.

If the noise blanking period is too short, then noise impulses will not be blanked effectively. On the other hand, if the noise blanking period is too long, then it will impair the decoder's ability to decode high speed Morse. For example, 24wpm Morse has dits lasting 50 milliseconds.



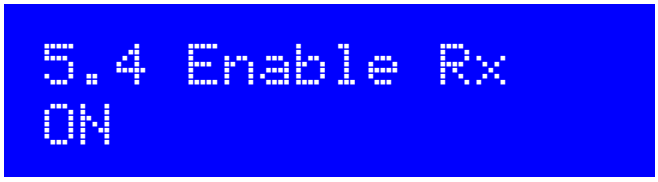
The duration of dits and dahs is measured in order to define a threshold at which to define a tone burst as a dit or a dah, and whether to define no tone at all as an inter-symbol, inter-character or inter-word gap. The measurement of this timing is implemented via an exponential moving average, whose averaging duration is determined by this parameter (the weight of each new measured symbol in the accumulated average).

If the exponential moving average is too fast (the parameter value is too low) then noise etc will throw off the timing averages too easily. If the exponential moving average is too slow (the parameter value is too high), then too many characters of the other station's transmission will be missed, while we try to adjust to the speed of his sending. This can be particularly offensive in some contest or pileup situations where exchanges are very short.

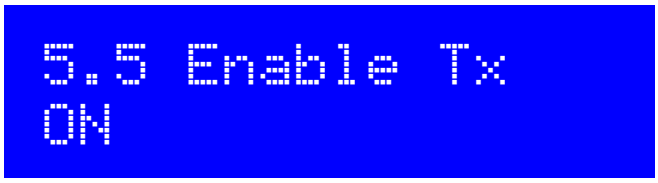


The decoder maintains an amplitude threshold, which it uses to decide whether a tone is detected or not. The level of this threshold must be varied automatically in order to cope with stations having a wide range of different signal strengths. Other perils may include QSB (signal fading) of the station you are listening to. The amplitude threshold is implemented via an exponential moving average. The weighting of each new sample (every 4ms) added to the accumulated exponential moving average value is the reciprocal of this parameter.

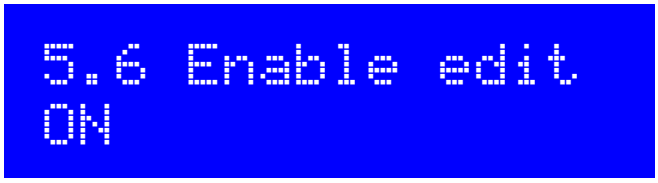
If the exponential moving average is too fast (the parameter value is too low) then noise etc will too easily throw off the amplitude threshold and it may take time to recover to its proper level. If the exponential moving average is too slow (the parameter value is too high), then it may take too long to adjust to the received station's amplitude, resulting in missed characters while the decoder slowly adjusts itself. It would also be too slow to automatically respond to QSB (signal fading).



An experienced CW operator may well dislike the CW decoder scrolling across the display all the time. With this setting you can switch "Enable RX decode" to OFF and the receiver decoding is disabled.



With this setting you can switch "Enable TX decode" to OFF and the transmit decoding is disabled. When this setting is ON, the CW decoder will decode your own keying and display it on the screen while you transmit. For an experienced CW operator that may be distracting too!



This parameter enables CW decoding while editing. When ON, anything you key during editing of NUMBER or TEXT type configuration parameters, edits the parameter. This is a really useful feature that makes it very easy to enter frequencies or stored messages, for example.

4.24 Beacon menu

The beacon function is an added bonus feature of this QRP Labs CW transceiver kit! We already have extensive experience for several years, developing the Ultimate-series QRSS/WSPR transmitter kits (current incarnation, the Utimate3S). These have a huge array of functionality and modes including CW, QRSS, DFCW, FSKCW, Hellscreiber (full speed and slow FSK), WSPR, JT9, JT65, ISCAT, Opera and PI4. The vast majority of people use the Ultimate3S kit for WSPR operation. Since it costs nothing (no extra hardware, at least) to add this functionality to the CW transceiver, why not! Let's do it!

The CW transceiver beacon function therefore contains a simplified WSPR implementation which can transmit standard WSPR messages. It also has a GPS interface for discipline of time, frequency and Maidenhead locator. The implementation of course does not have the full range of flexibility and functionality as the Ultimate3S kit.

WARNING: WSPR transmissions operate a continuous 100% key-down duty-cycle for almost 2 minutes. You should check carefully whether the BS170's get too hot during this period. WSPR is much more demanding on the PA transistors than CW. If you are operating with higher output powers (by using a relatively high power supply voltage) then you might notice the BS170s get unreasonably hot. Even though the output stage is Class-E, nothing is 100% efficient! It will still dissipate some power as heat. In this case you may need to fix the BS170 transistors to a small heatsink to help dissipate the heat. The BS170 transistors Q1, Q2 and Q3 are placed at the edge of the PCB with their flat sides facing outwards, to facilitate this

The beacon function can also operate a CW beacon.

Weak Signal Propagation Reporter

WSPR stands for Weak Signal Propagation Reporter. It is digital message format filled with clever forward error correction. The message consists of three parts: the operator's callsign, Maidenhead locator (4-character, e.g. IO90) and two digits specifying the power. At the receiving station, messages are decoded and uploaded to a central internet database. At any time you can go to WSPRnet http://wsprnet.org and click on the map, enter your callsign (and other filters if you wish), and see a map of where your signal is being heard.

You could also undertake more in-depth propagation studies by downloading the database of reception reports.

The WSPR message is encoded into a set of 162 symbols, each may be 0, 1, 2 or 3, using a compressed data format with forward error correction. The symbols are transmitted as tones, each tone separated by 12,000 / 8,192 Hz i.e. approximately 1.46Hz. The duration of each symbol is the reciprocal of the tone spacing, which is approximately 0.683 seconds. WSPR messages take about 110.6 seconds to transmit, and always start at even minutes past the hour.

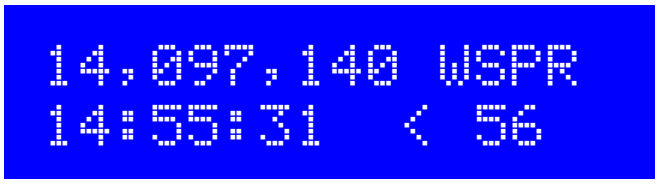
Due to the very narrow 6Hz bandwidth of the transmission, and the clever forward error correction, WSPR signals can propagate globally even with a fraction of a watt.

In WSPR, timing is critical, so when using WSPR you must set the time configuration parameter as exactly as possible. Be sure to keep the editing cursor under the rightmost (1-minute) digit of the time parameter, watch your clock until the seconds turn over to 00, and then press the left button. This will ensure the seconds are in sync with your real clock time. If careful attention is given to setting the frequency and the real time clock, then successful WSPR reports will be obtained. Of course these things are easier if you are using a GPS module: the Maidenhead

locator will be calculated from the received latitude and longitude, and the time decoded nicely from the GPS serial data stream.

The microcontroller in this kit takes care of the WSPR message encoding algorithm, without any assistance from a PC host computer. It also calculates the tone spacing and symbol duration.

In between message transmissions, the display will show instead just a clock (see below), while we wait patiently for the next WSPR transmission to begin, according to the settings of the configuration parameters Frame and Start. This is useful for checking that the time on your kit is accurately set. The display also shows the minute at which the next frame will start transmitting. In the example below, the time is 14:55:31 UT and the next frame will start at 14:56:01.

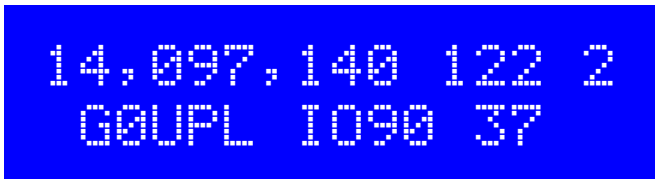


When a GPS unit is connected, the firmware automatically uses the 1 pulse-per-second signal to measure the transmit frequency and compensate for any inaccuracy due to calibration error or frequency drift due to temperature. The serial data stream from the GPS is used to set the real time clock (for syncing the WSPR transmission timing). The Maidenhead locator is computed from the latitude and longitude information parsed from the GPS serial data.

A WSPR transmission takes 1 minute and 52 seconds. The GPS time and location data is parsed from the GPS serial data stream at the END of every WSPR transmission. On the first WSPR transmission after power up, the time may not be accurate. The 27MHz synthesiser reference frequency is measured and adjusted, in a 4 second period after the end of the WSPR transmission. Parsing the time and location, and adjusting the 27MHz reference frequency, all takes place in under 7 seconds. The process is completed before the start of the next WSPR transmission slot. Nevertheless, you should not configure your kit for continuous WSPR transmissions in every 2-minute WSPR slot (Frame parameter is 2), which is considered very antisocial to fellow WSPR operators.

A GPS receiver isn't essential for WSPR operation but it is highly recommended because it makes operation more accurate, easy and fun.

During the actual WSPR message transmission, the display shows something like this:

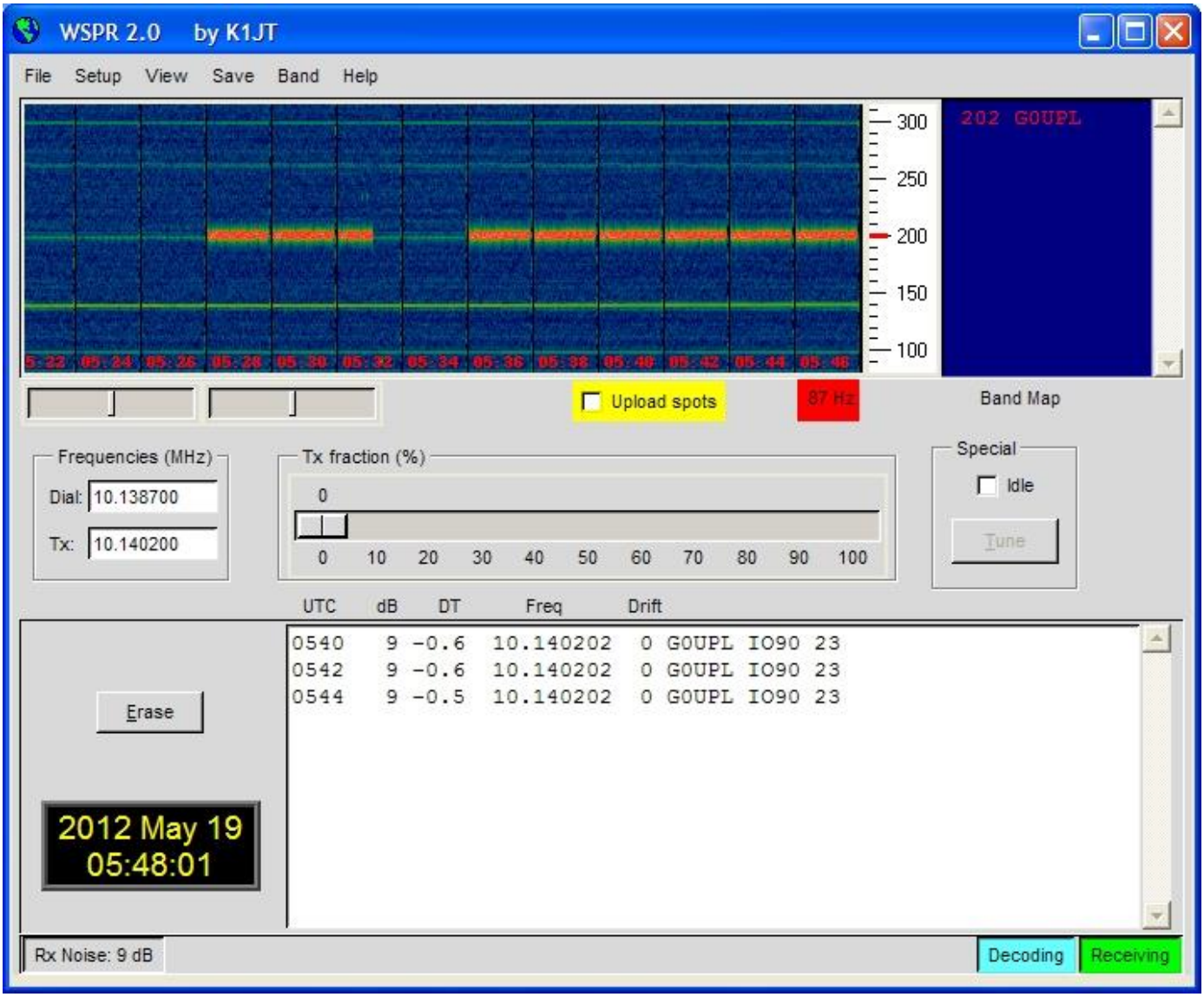


The elements of this display are explained as follows:

- 14,097,140 Transmission frequency (tone 0)
- 122 Current symbol is 122 (of 162 symbols making up a WSPR transmission)
- 2 Current tone being transmitted (one of 0, 1, 2 or 3)
- G0UPL Callsign encoded into the transmission
- IO90 5-character Maidenhead Locator square, encoded into the transmission
- 37 Power in dBm, encoded into the transmission

On a PC spectrum display such as the Argo software http://www.weaksignals.com/ WSPR messages look something like the screenshot below when received locally (or usually worse, because you probably are over-driving your receiver when receiving your own signal!):

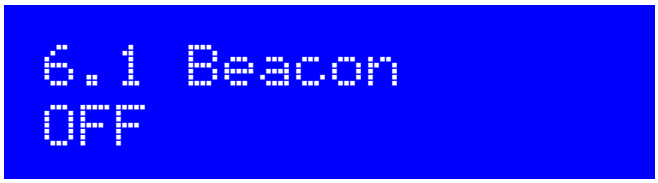
WSPR decoding takes place in the WSPR program by K1JT (see http://physics.princeton.edu/pulsar/K1JT/wspr.html). Below is a screenshot showing the WSPR 2.0 screen following reception of a few transmissions (output frequency = 1,500Hz, Frame = 02, Start = 00).



A number of other software packages are now also capable of decoding WSPR transmissions.

NOTE: the GPS interface is connected in parallel in the circuit, with the paddle. Therefore, you should NOT have the GPS connected, while operating the radio as an ordinary CW transceiver. If you do, the GPS serial data and 1pps will key the transmitter! Disconnect the GPS before using the radio as a CW transceiver.

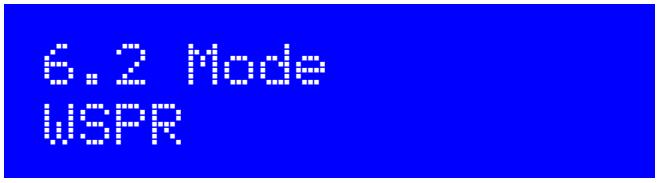
The following sections describe the configuration parameters in the Beacon menu.



This is the master switch controlling whether beacon mode is enabled or not. If beacon mode is enabled, then the radio starts operating in beacon mode on power-up.

While beacon mode is operational, it can be cancelled immediately at any time by pressing the right (exit) button. Remember that you should not have a GPS connected, when in ordinary CW transceiver mode – it would key the transmitter since the GPS and paddle share the same processor I/O signals.

Beacon mode is entered upon leaving the configuration menu system, if the beacon is enabled by having this parameter set to ON.



This parameter determines the transmission mode during beacon operation. There are two possible beacon modes:

CW: the kit simply sends stored message 1 repeatedly, according to the configured Keyer speed, and with message start timing determined by the Frame and Start parameters (see next sections).

WSPR: the kit sends WSPR according to the configuration parameters in the following sections.

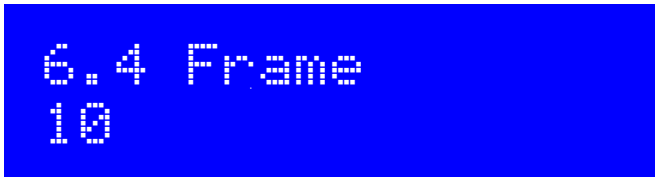
This parameter determines the transmission frequency during beacon operation. In WSPR mode, this is the frequency of tone 0.

It should be noted that the WSPR sub-bands on the bands are only 200Hz wide. You need to have your 27MHz reference oscillator accurately calibrated so that you can be sure your actual transmission frequency is close to the frequency specified in this parameter. You also need to specify the correct frequency so that your transmission is inside the appropriate 200Hz sub-band.

Note that these are the actual transmission frequencies, there is no CW offset, no RIT, no other modifications. The specified frequency is also DIFFERENT from the “USB dial frequency” specified at WSPRnet http://wsprnet.org – the “USB dial frequency” is 1500Hz lower than the actual transmit frequency, so that the decoded audio is at 1500Hz.

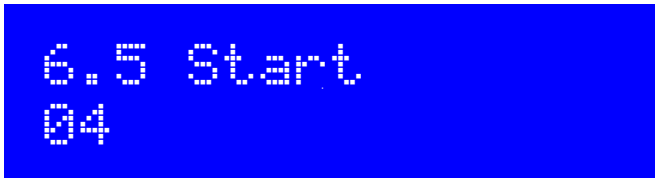
Therefore, ensure that for WPSR transmissions, you choose a frequency in one of the WSPR sub-bands according to the following table:

80m:	3.594000 – 3.594200
60m:	5.288600 – 5.288800
40m:	7.040000 – 7.040200
30m:	10.140100 – 10.140300
20m:	14.097000 – 14.097200
17m:	18.106000 – 18.106200
15m:	21.096000 – 21.096200
12m:	24.926000 – 24.926200
10m:	28.126000 – 28.126200



This parameter defines the repetition rate of the WSPR transmission. In the example shown here, Frame 10, this means that the WSPR message will be transmitted once every 10 minutes.

Transmission in every 2 minute WSPR slot is considered anti-social. 10 minute repeat transmissions is usually considered normal.



If everybody transmits with 10 minute repetition rate starting on the hour, then there will be bursts of activity every 10 minutes where everyone is transmitting at once, and the potential for interference from another station will be large. To avoid this, you can define the start timer. In this example a start time of 04 means that the first transmission will start at 4 minutes past the hour, and subsequent transmissions will commence at Frame minutes after that – in this case, 14, 24, 34 etc minutes past the hour.

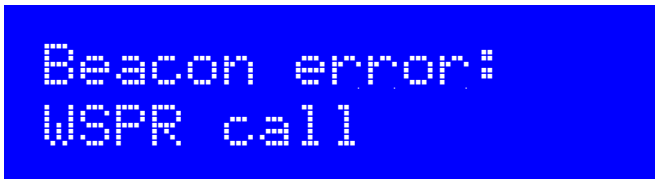
The WSPR callsign is the first parameter which is encoded into the WSPR message. Your callsign must obey certain restrictions imposed by the WSPR protocol. These restrictions helps ensure that the WSPR encoding process can compress callsigns efficiently, along with the Maidenhead Locator square and Power level, into only 50 binary bits of information in total.

The callsign can only be 4 to 6 characters long. The callsign must consist of the following:

1) One character which can be A-Z or 0-9 or a SPACE
2) One character which can be A-Z or 0-9
3) One character which must be a number 0-9
4) Three characters which can be A-Z or a SPACE

For callsigns such as mine, consisting of 5 characters, I must enter a space character as the first character in order to satisfy these callsign rules.

If the callsign you enter does not obey the necessary rules, then an error message is displayed on exiting the configuration menu system:

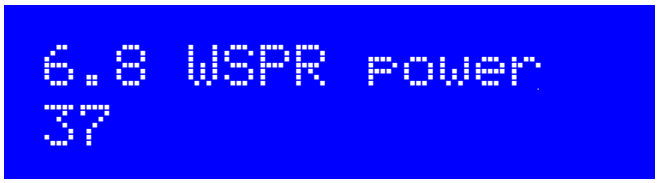


In this case, go back to the WSPR Call parameter and try to understand how to correct it in order to make your callsign satisfy the requirements.

Note that if entering the callsign text using the key, you cannot enter a space with the key! So, you will need to enter the initial space character (if required) using the buttons and rotary encoder as discussed in the earlier section on editing TEXT parameters.

The Locator is the second parameter which is encoded into the WSPR message. It is the 4-character Maidenhead square. The text you enter here, must be a valid Maidenhead square, otherwise an error message will be generated on exiting the configuration menu system.

If you have connected a GPS receiver, the GPS receiver can update the Locator, computing it from the latitude and longitude information contained in the serial data string from the GPS receiver module.



The third and final parameter encoded into the WSPR message is the transmitter power, defined in dBm. Note that this parameter is manually edited here and is encoded into the WSPR message. It does NOT indicate a measured power which is actually transmitted. This is a common misconception. It only indicates what number the operator configured for the WSPR message encoding.

WSPR powers are constrained to certain values 0, 3, 7, 10, 13, 17, 20, 23, 27, 30, 33, 37, 40, 43, 47, 50, 53, 57 and 60dBm. If you specify a value not in this list, then an error message will be generated on exiting the configuration menu system.

In this example, the configured value is 37dBm which corresponds to 5 watts of RF transmitter output.

The kit has an internal real time clock which is used for timing the beacon (WSPR and CW) transmissions. This configuration allows you to set the real time clock. The clock setting takes effect, and the seconds (internally counted but not displayed) are set to zero, when editing of this menu is concluded: either by pressing the left button after setting the minutes, or by pressing the right button at any time.

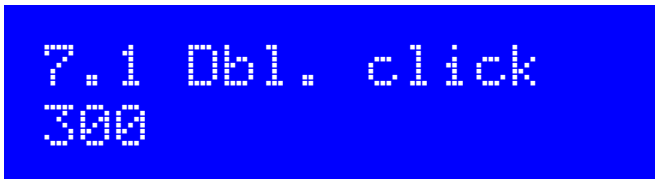
Accurate timing is critical for the WSPR (etc) modes. With careful adjustment of the “System frq.” parameter, the 20MHz oscillator will keep accurate enough time for WSPR for up to a week, even without GPS.

Note that if using a GPS module and if there is a valid serial data stream properly decoded, then the time is set automatically.

An accurate real time clock is not required other than for the WSPR beacon transmission timing.

4.25 Other menu

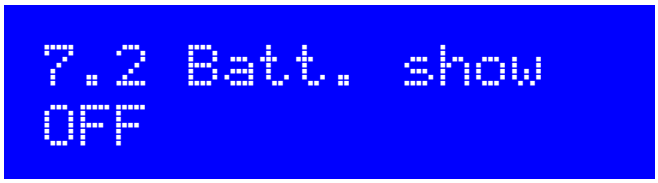
This menu contains miscellaneous other settings which do not fit well in other configuration menu categories.



This is a NUMBER parameter which controls the decisions on what type of press has been made to a button. By default, it is set to 300 milliseconds (as shown here) but you may alter this if you wish.

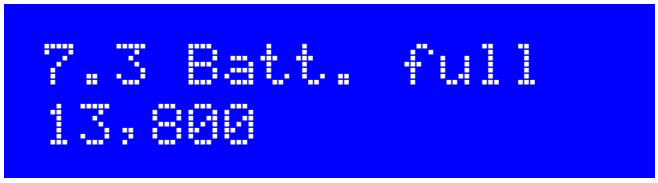
It is the number of milliseconds after first pressing the button, at which certain decisions are made:

a) If you have not pressed the button again after this interval, then it means you intended a single press.
b) If you are STILL pressing the button all this time later, 300 milliseconds after the first press, then it means you executed a “single long press”.
c) If you pressed the button again before the 300 milliseconds elapsed, it is a “double click”.

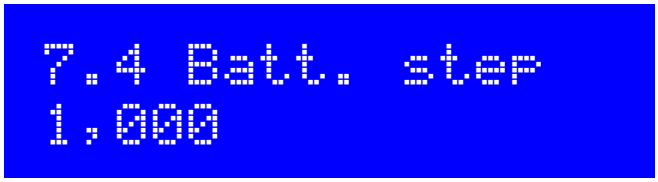


This BOOLEAN parameter controls whether or not the battery icon is displayed on the screen at the top right corner. Measurement and display of battery voltage may be useful to those operators who intend to operate the radio from battery power, for example during portable operations.

In order to measure and display the battery voltage, it is necessary to jumper the “DVM” input header pin to the supply voltage header pin. This hardware wiring is described elsewhere in this document.



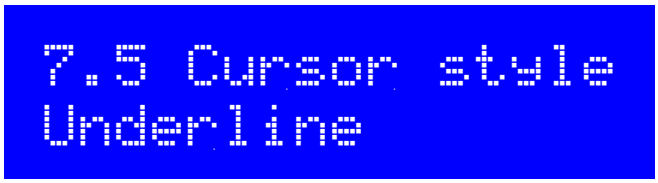
This is the voltage, specified in millivolts, at which the battery is considered “full”.



This is the step, specified in millivolts, for each bar of the battery icon.

In this example, Batt. full is defined as 13.8V and the step is 1V. The battery icon has 7 possible states, ranging from empty to full and 5 intermediate states in between. The meaning of the displayed icon will be, in this example:

- Full: 12.81V to 13.8V (and indeed, above 13.8V also)
- 5 bars: 11.81V to 12.8V
- 4 bars: 10.81V to 11.8V
- 3 bars: 9.81V to 10.8V
- 2 bars: 8.81V to 9.8V
- 1 bar: 7.81V to 8.8V
- Empty: 7.8V and below

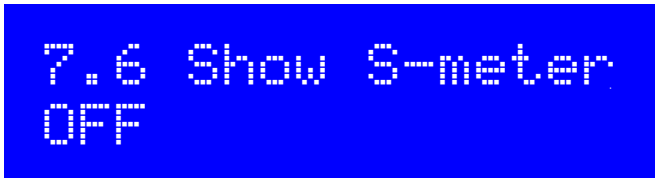


Two different cursor styles are possible. You can select your favourite, here. The two possible values are:

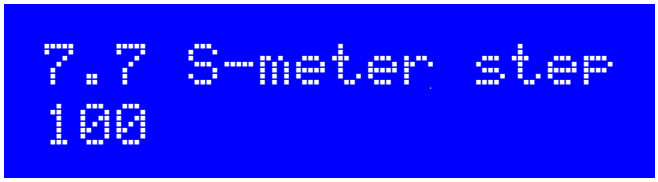
Underline: a simple solid line under the character being edited

Blink: the display alternates between the edited character, and a solid white block

Note that this setting only affects the cursor that is shown during menu system editing. In normal operating mode, the underline cursor is always used for tuning rate indication, regardless of the Cursor style setting.



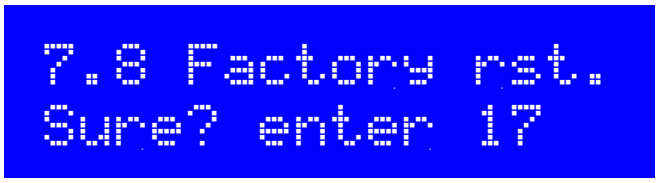
A rudimentary S-meter is shown in the top right of the display, if this parameter is set to ON.



This NUMBER parameter effectively defines the sensitivity of the S-meter. The S meter has 12 bars maximum. Each bar of the S-meter equates to the specified “S-meter step” number on the amplitude measurement scale. So, in this case the scale of the 12 bars of the S-meter will mean 0 to 1200 in amplitude measurement terms.

A lower S-meter step number makes the S-meter more sensitive; a higher number makes it LESS sensitive.

The S-meter reading is indicative only, it does not have any highly accurate meaning. In any case the audio sampled by the Analogue to Digital converter is after the audio amplifier gain stages so the amplitude depends on the receiver gain setting.



This menu item can be used to cause a factory reset. Factory reset returns you radio to the supplied default factory configuration. Everything is erased and set back to the default parameter values.

In order to prevent accidentally triggering this drastic step, the factory reset is implemented as a virtual 2-digit numeric parameter. To cause a reset:

- Press the left (“select”) button to start editing this menu item. 00 is displayed
- Use the rotary encoder to change the value to 17
- Press the left (“select”) or right (“back”) buttons to enter the value 17, which causes the factory reset

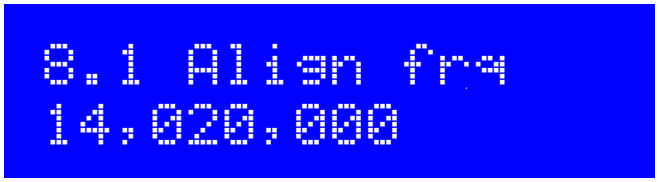
Factory reset takes a few seconds while the entire EEPROM contents are written. After the factory reset you will see the “Select band” prompt again, the same as when you first power up the radio.

4.26 Alignment menu

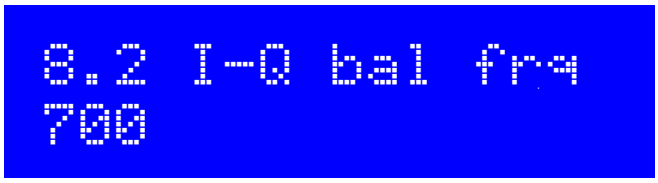
The alignment menu contains several tools for alignment and calibration of the radio, using the built-in signal generator to inject a test signal back into the receiver’s front end.

Correct use of these alignment tools makes it possible to adjust and align the receiver without any external test equipment at all.

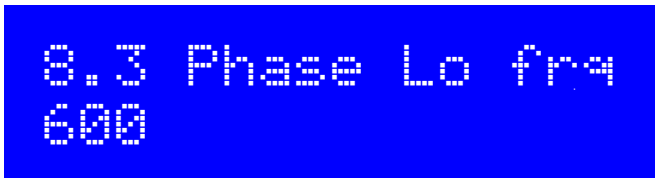
Please also refer to the alignment/adjustment section at the end of the assembly part of the manual, which describes how to use these tools to peak the Band Pass Filter response, and to minimise the unwanted sideband using the I-Q balance and audio phase shift adjustments.



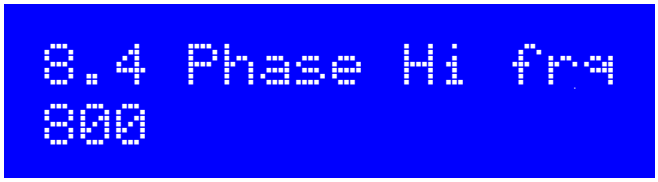
The Align frq. parameter specifies the frequency which the alignment tools operate. You should ensure that the frequency specified is in the centre of the CW section of the band of operation of your radio.



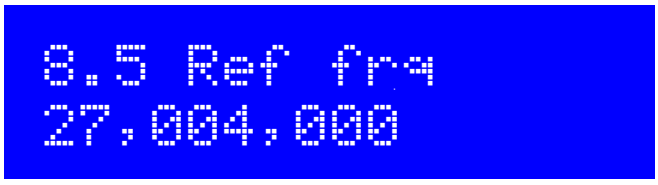
The audio frequency at which the I-Q balance adjustment is performed. By default this is set to 700Hz, the centre of the audio filter and what is usually used for the CW offset frequency.



The audio frequency at which the LOW audio frequency 90-degree phase shift is performed.



The audio frequency at which the HIGH audio frequency 90-degree phase shift is performed.

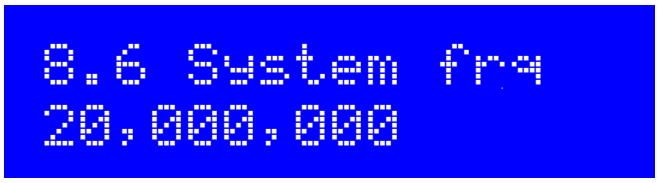


The reference frequency for the Si5351A synthesiser chip. If you set this to the actual oscillation frequency of the 27MHz crystal, then the output frequency of the radio will be accurate.

Usually the 27MHz crystals oscillate between 3 to 5kHz too high. This is the reason for the default setting of 27.004MHz (4kHz high).

You can measure the 27MHz reference value yourself if you have a reliable general coverage receiver whose frequency is known to be accurate. There are other methods too! You could set the signal generator to a certain frequency such as 10MHz. Then beat it against a 10MHz reference, for example. Measurement of the audio offset would enable a calculation of what the correct 27MHz reference value ought to be.

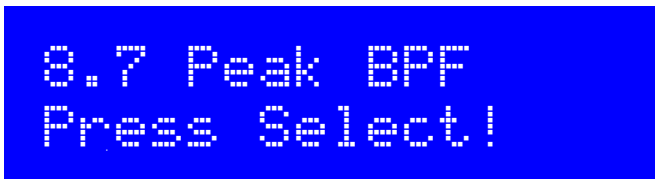
If a GPS unit is available, then this makes everything very simple – then you can use the GPS calibration tools to calibrate this value (see below).



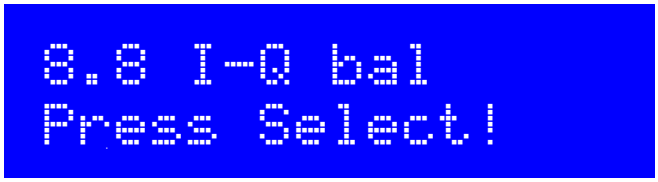
The system oscillator frequency. In this radio, the microcontroller clock is a 20MHz crystal. If you are able to measure the actual 20MHz frequency (without disturbing it), you can enter the frequency here. You could, for example, listen for the 20MHz radiation on a general coverage receiver whose frequency indication you trust.

The 20MHz system clock is used for timing purposes only. It is not very critical, unless you intend to operate the WSPR beacon function without a GPS.

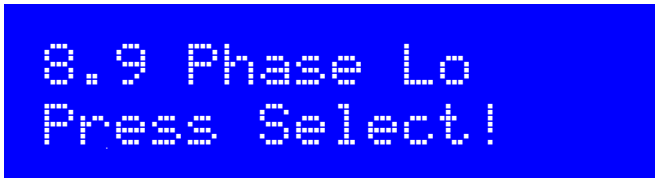
If a GPS unit is available, then you can use the GPS calibration tools to calibrate this value (see below).



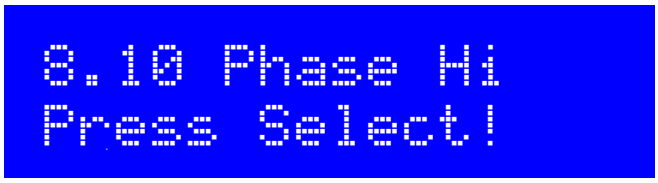
Pressing the left (select) button activates the signal generator on the configured frequency and displays a bar on the screen showing the measured audio amplitude at 700Hz. The Band Pass Filter trimmer capacitor should be adjusted for maximum measured amplitude. This process is described in the earlier section “Adjustment and alignment” at the end of the assembly part of the manual.



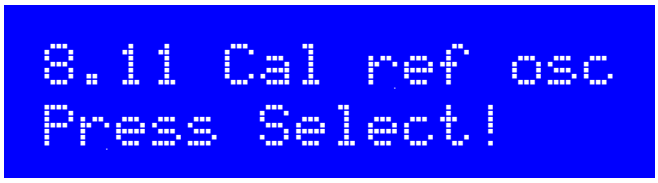
Pressing the left (select) button activates the signal generator on the configured frequency with an offset to put the received signal in the unwanted sideband. It displays a bar on the screen showing the measured audio amplitude at 700Hz. The I-Q balance trimmer potentiometer should be adjusted for minimum measured amplitude. This process is described in the earlier section “Adjustment and alignment” at the end of the assembly part of the manual.



Pressing the left (select) button activates the signal generator on the configured frequency with an offset to put the received signal in the unwanted sideband at the audio frequency configured (for example, 600Hz). It displays a bar on the screen showing the measured audio amplitude. The Low frequency phase adjustment trimmer potentiometer should be adjusted for minimum measured amplitude. This process is described in the earlier section “Adjustment and alignment” at the end of the assembly part of the manual.



Pressing the left (select) button activates the signal generator on the configured frequency with an offset to put the received signal in the unwanted sideband at the audio frequency configured (for example, 800Hz). It displays a bar on the screen showing the measured audio amplitude. The High frequency phase adjustment trimmer potentiometer should be adjusted for minimum measured amplitude. This process is described in the earlier section “Adjustment and alignment” at the end of the assembly part of the manual.

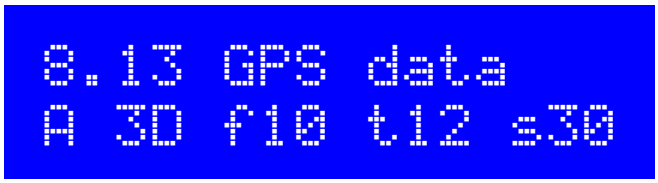


If you have connected a GPS receiver module such as the QRP Labs QLG1 GPS receiver http://qrp-labs.com/qlg1 with a one pulse per second (1pps) output signal, then when you press the left (select) button the microcontroller uses the 1pps as a frequency counter gate. The 27MHz reference value is divided by 4 and routed to the Timer1 input of the microcontroller, where it is counted for 4 seconds to obtain a frequency count. This value is stored in the EEPROM in the configuration parameter editable in menu item “8.5 Ref frq”.

The firmware calculates compensation adjustments to correct for the frequency error and ensure an output frequency precisely on target (typically with sub Hz accuracy). Naturally changes in ambient temperature will also cause the crystal oscillator frequency to change somewhat.

Using the GPS to calibrate the reference oscillator, system oscillator, or GPS info display (see later section), also automatically parses the serial data stream and sets the real-time clock and the Maidenhead Locator square (computed from latitude and longitude); these are used by the Beacon mode when using the radio as a WSPR beacon.

If you have connected a GPS receiver module such as the QRP Labs QLG1 GPS receiver http://qrp-labs.com/qlg1 with a one pulse per second (1pps) output signal, then when you press the left (select) button the microcontroller uses the 1pps as a frequency counter gate. It counts the 20MHz system clock internally for 4 seconds to obtain a frequency count. This value is stored in the EEPROM in the configuration parameter editable in menu item “8.6 System frq”.



When a GPS receiver (1pps and serial data) is connected and the alignment/calibration functions are enabled by pressing the left (select) button, data from the GPS is shown on the display. The five displayed info fields are to be interpreted as follows:

- A valid GPS Data (V would mean invalid data)
- 3D 3D fix (could be None, 2D or 3D)
- f10 GPS solution fix uses 10 satellites
- t12 GPS receiver is tracking 12 satellites
- s30 average signal strength of the 12 tracked satellites is 30.

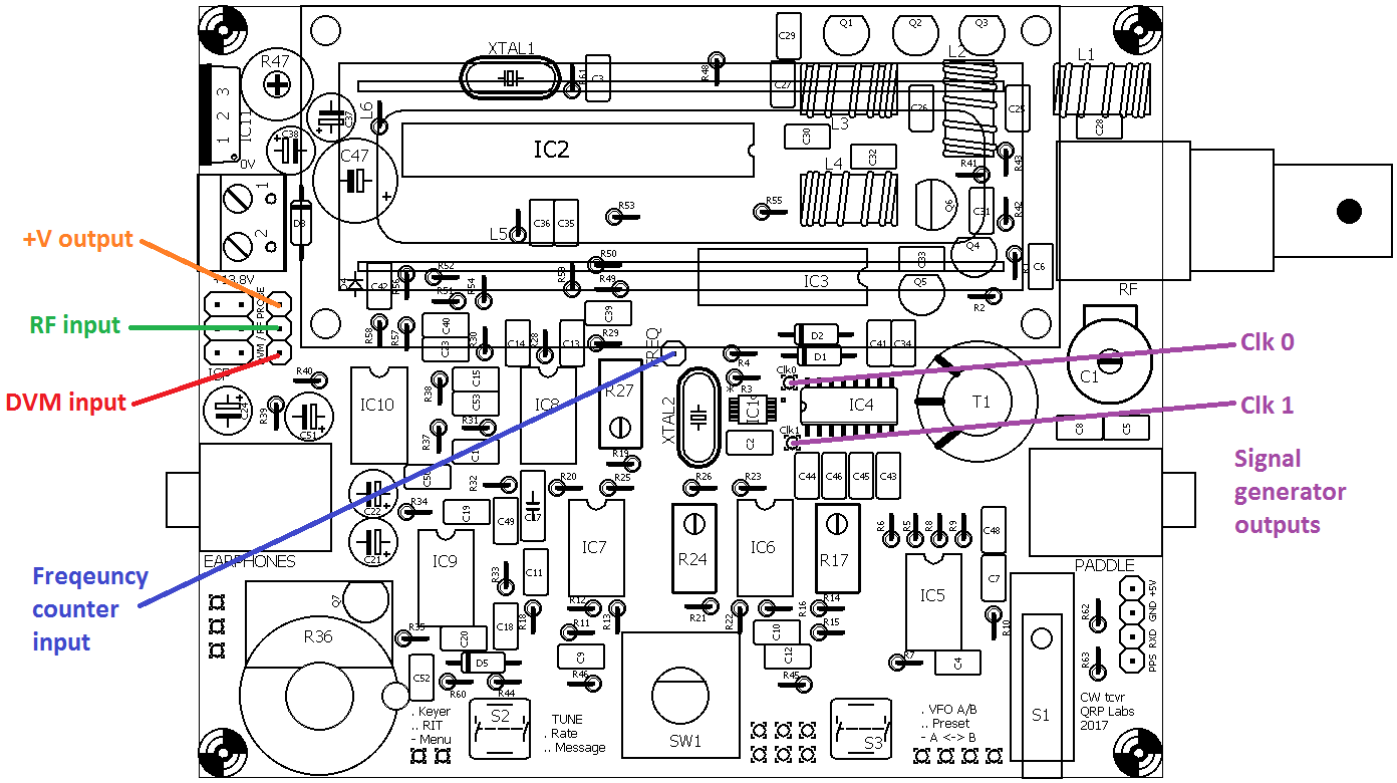
The GPS data menu item is for display of GPS information only. It can be useful to verify that the GPS is working properly and determine the best location for the GPS antenna, based on the number of tracked satellites and signal strength.

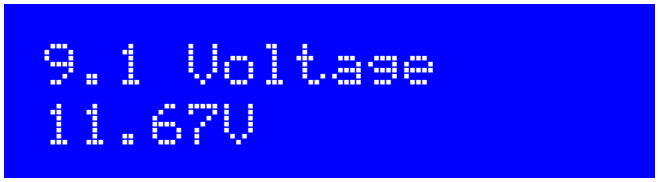
4.27 Test equipment

The test equipment menu provides several items of test equipment, which can be used to diagnose faults in your radio, and even for other testing purposes too!

The diagram below shows the connection points for the test equipment. **If you are attempting to measure external equipment or circuits, then you will also need to make a ground connection between the CW transceiver and the external circuit.**

These items of test equipment are described below. All except for the signal generator, are display-only items, with no corresponding stored configuration parameter.

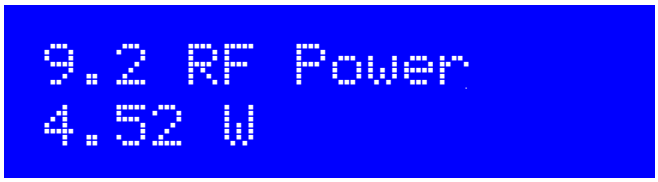




This the Digital Voltmeter (DVM) function. The measurable voltage range is 0-20V DC.

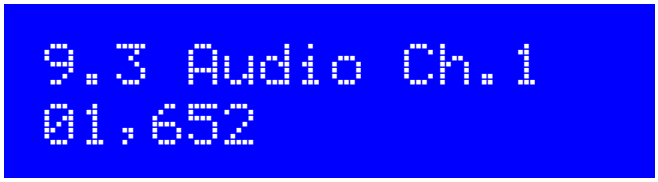
When the on-screen battery voltage display is used, you must connect the DVM input pin to the +V output pin (refer to previous diagram showing these connections).

In this example, a 12V power supply is being used to power the radio. The DVM input pin is jumpered to the +V output pin shown. The measured voltage of 11.67V is 12V minus the voltage drop over the reverse polarity protection diode D3.

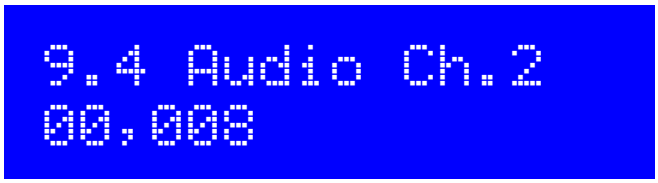


This the RF Power meter function. The measurable power range is 0-5W. The RF power measurement circuit uses a simple diode detector. There are non-linearities in the diode function, and differences from diode to diode; therefore, the measured power output is not highly accurate. However, it is certainly a very good indication and debugging tool.

Note that the RF Power meter and the DVM use the SAME Analogue to Digital (ADC) microcontroller input. You must not try to use the RF Power meter and the DVM at the same time! Ensure only one or the other is connected.



This item just displays the raw amplitude number calculated by the Goertzel algorithm for the audio channel 1, which is connected to the output of the audio amplifier. The value shown here will depend on the input signal level but also on the gain control. You can easily verify that as you turn up the gain control knob, the displayed value increases.



This item just displays the raw amplitude number calculated by the Goertzel algorithm for the audio channel 2, which is connected to the output of the phase nulling circuit. The value shown here is independent of the gain control. The value is much less than that measured on Audio Ch.1. This is because one ADC least significant bit equates to a voltage of about 5mV. The signal amplitude at that point in the receiver signal path is too low to produce much output from the ADC unless the received signal is very strong.

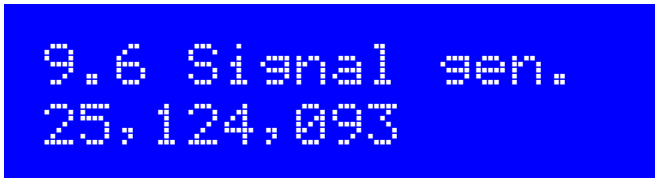
This is the frequency counter. The resolution is 10Hz, which results in an update rate of around 8 times per second approximately.

The frequency counter input is connected directly to the Timer1 input of the microcontroller. There is no pre-amplifier or level adjustment. The ATmega328P processor operates from a 5V supply. It expects the frequency counter input signals to be of an appropriate amplitude.

According to the ATmega328P datasheet, a “low” has a maximum input voltage of 1.5V and a “high” has a minimum input voltage of 3.5V. Therefore, the minimum amplitude signal to drive the frequency counter successfully would be 2V peak-peak centred on a 2.5V DC offset. Exercise great care to ensure that the voltage levels are correct, to avoid damaging the ATmega328P input pin.

Additionally, since the ATmega328P has synchronous timer inputs, the highest frequency which can be counted is limited to about 40% of the system clock (20MHz).

Therefore, this frequency counter has range of 0 to 8MHz. The upper frequency end of this range will only be achieved in the case of a nice tidy squarewave of adequate amplitude an 50% duty cycle. Lower quality signals will result in a lower upper frequency limit.



The signal generator function is switched off until you press the left button to select it. Then the cursor appears under the 10MHz digit. You can then adjust this number just like any other configuration parameter. Use the rotary encoder to increase or decrease the frequency. Press-and hold, then turn the rotary encoder, to change the tune rate (move the cursor left or right).

As you adjust the frequency, the output frequency of the synthesiser is tuned to match.

The output frequency range is from 3.5kHz to 200MHz. The Si5351A Synthesiser chip datasheet maximum frequency specification is 200MHz. However, in practice it has been found to still operate reliably up to almost 300MHz.

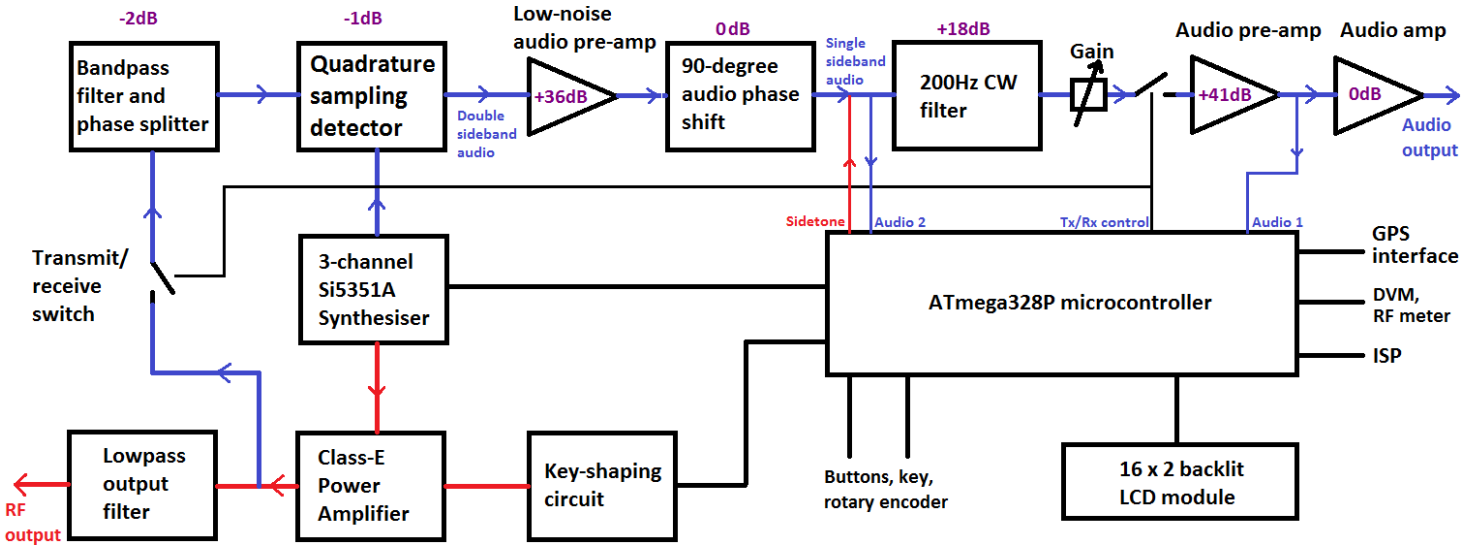
The output waveform is a squarewave with amplitude about 3.5V peak-peak. The output is at the Clk0 and Clk1 terminals (see diagram, above).

Note that the output is directly connected to the Si5351A Synthesiser chip's output pin. Take very great care not to short the output to ground or in any other way abuse this output! If you damage the Si5351A chip it is very difficult to replace

Note that where possible, the Si5351A's Clk0 and Clk1 outputs are operated in quadrature with a 90-degree phase offset. The relationship does not hold at frequencies below 3.2MHz and may not hold into VHF (To Be Determined).

5. Circuit design of the QCX

5.1 Block diagram and summary



This CW transceiver is a high performance, yet simple and low cost, analogue design. The transmitter uses a high efficiency Class-E amplifier which results in low current draw on transmit, and inexpensive transistors with no heatsinks.

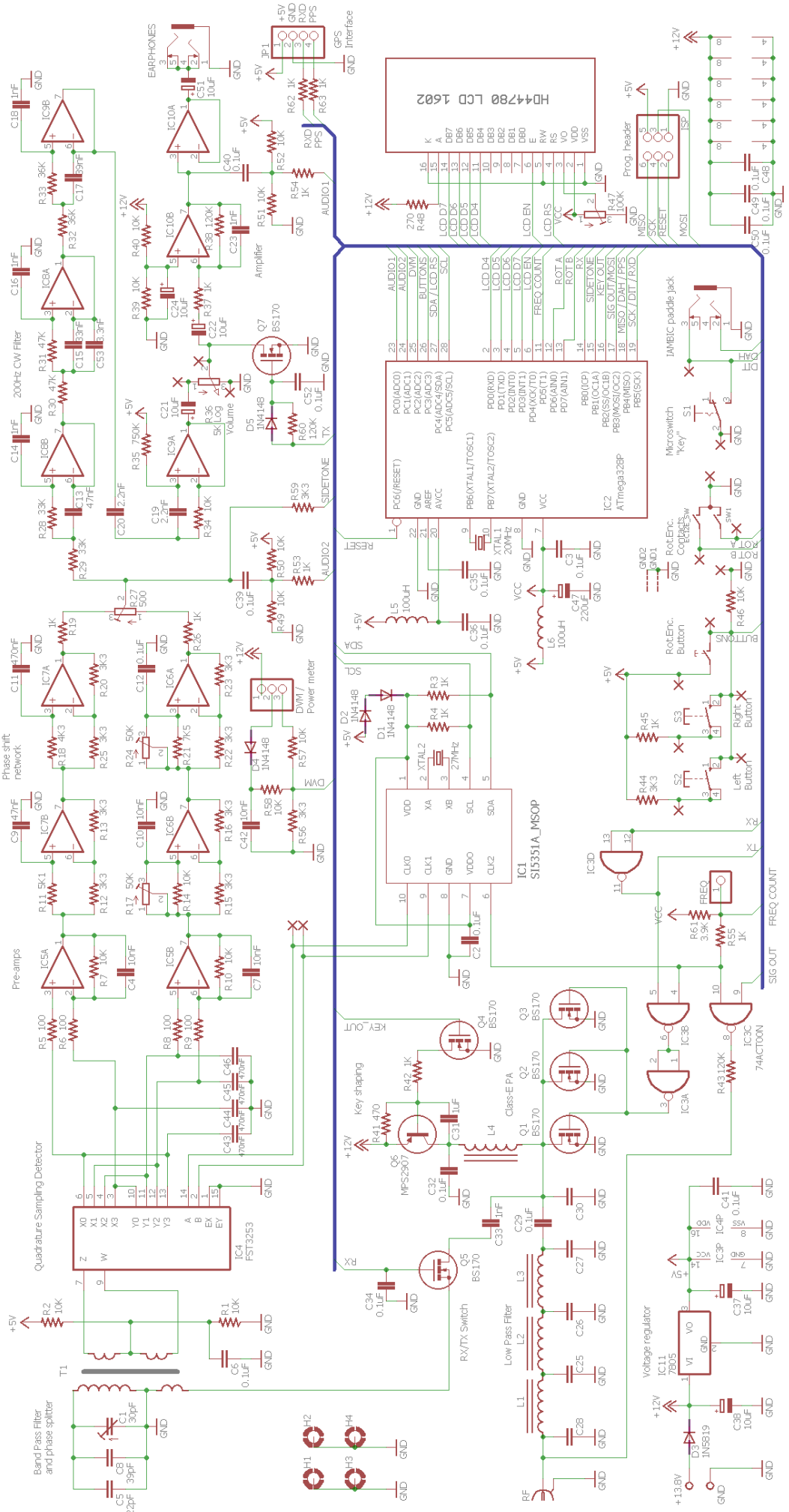
The receiver is a direct conversion type utilising the famous high performance Quadrature Sampling Detector, also known sometimes as the “Tayloe Detector” or even “I-Q Mixer”. This receiver front end architecture has been used in the early Flex Software Defined Radios, Softrock series, Norcal NC2030 and many other SDR’s and other high performance front ends. The detector has very high third order intercept (IP3) and dynamic range, as well as low loss.

The resulting I & Q outputs are at audio baseband and go through a 90-degree phase shift network which cancels the unwanted sideband. A 200Hz bandwidth CW filter is followed by more amplification and drives common earphones.

The oscillators in the transceiver are provided by the modern Si5351A digital phase locked loop IC controlled by the microcontroller.

Permeating the entire design is microprocessor control by the ATmega328P microcontroller. This allows implementation of a large number of functions normally only found in radios costing 10-100x the price!

A really nice feature of the design is the built-in alignment and test equipment, which make it possible to build, align and even debug the assembly of the radio, all with NO additional test equipment.


HD44780 LCD 1602
ATmega328P
SI5351A_MSOP
FST3253
74ACT00N
Quadrature Sampling Detector
Pre-amps
Phase shift network
Band Pass Filter and phase splitter
Key shaping
Class-E PA
Low Pass Filter
RX/TX Switch
Voltage regulator
Amplifier
200Hz CW Filter
Volume
DVM / Power meter
Rot.Enc. Button
Rot.Enc. Codetech sw
Microswitch "Key"
IAMBIC paddle jack
Prog. header
GPS Interface
EARPHONES
SIG OUT
FREQ COUNT

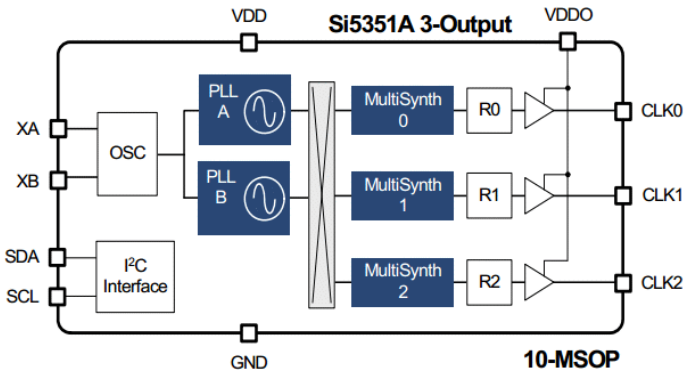
5.2 Circuit diagram

The complete circuit diagram (schematic) is shown on the previous page.

5.3 Synthesised oscillator

I always start with building the VFO of a radio. It was the hardest thing to get right. How to get that analogue LC-tuned VFO accurate, free of drift, free of chirp, tuning over the required range, and with mechanical gearing to be able to make fine frequency adjustments? A real challenge. Not any more! Now we have Direct Digital Synthesis (DDS) ICs and Digital Phase Locked Loop (PLL) ICs, inexpensive and easy to use, that solve all the problems.

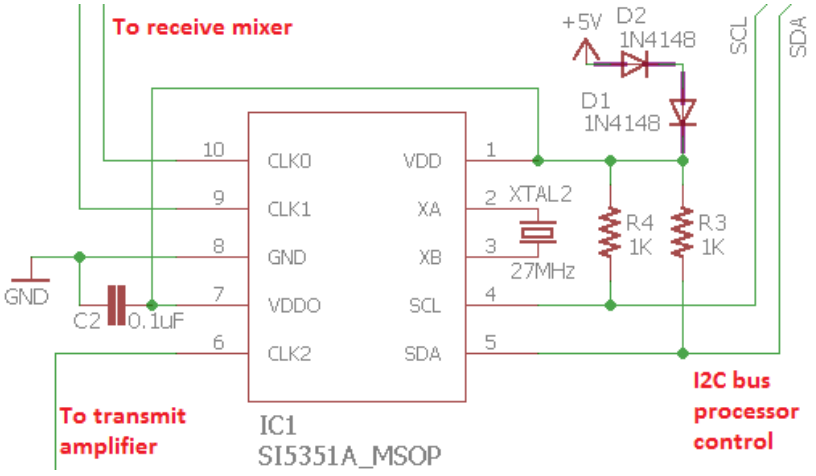
The Si5351A Synthesiser chip used in this design provides three separate frequency outputs, with a frequency range spanning 3.5kHz to 200MHz. The frequency stability is governed by the 27MHz crystal reference. Pretty stable, in other words.



The block diagram (right) is taken from the SiLabs Si5351A datasheet. Briefly, the 27MHz reference oscillator is multiplied up to an internal Voltage Controlled Oscillator in the range 600-900MHz (the PLL), then divided down to produce the final output frequency. The multiplication up and the division own are both fractional and so the frequency resolution is extremely finely controlled. The chip has two PLLs and three output divider units.

For best jitter performance, the Si5351A datasheet recommends the use of even integer dividers (no fractional component) in the MultiSynth dividers and in this CW transceiver design, this recommendation is followed.

The synthesiser section of the circuit diagram is shown here (right). The Si5351A datasheet dictates the use of a 25 or 27MHz crystal. QRP Labs has always used the 27MHz crystal in our designs because it allowed us to obtain precise 1.46Hz tone spacing for WSPR transmissions all the way up to the 2m amateur band (145MHz). Those calculations don't work out with the 25MHz crystal. This requirement doesn't apply to this CW transceiver design but economics of scale means there are advantages to sticking with the same component values, all other things being equal!



The Si5351A has a large number of internal 8-bit registers to control the synthesiser behaviour, and these are programmed by the microcontroller using the I2C serial protocol. 1K resistors R3 and R4 are pull-ups required for the operation of the bus at 400kHz.

The Si5351A chip requires a 3.0 to 3.6V supply (nominally 3.3V) but the rest of this transceiver's digital circuits operate with a 5V supply. For the reduction of complexity and costs, two 1N4148 diodes in series are used here to drop the 5V to a suitable voltage for the Si5351A. It works well.

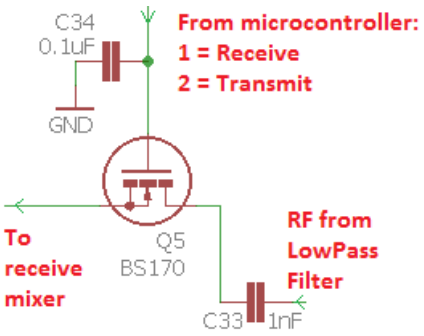
There are three outputs of the Si5351A synthesiser and these are all used to good advantage. The Clk2 output is used to feed the transmit power amplifier, and the Clk0/1 outputs are used to drive the Quadrature Sampling Detector (QSD) during receive. These outputs can be switched on and off under the command of the microcontroller. This provides an opportunity for some simplification because the Clk0/1 outputs can be simply switched off entirely during transmit. This relieves pressure on the transmit/receive switch. There just cannot be any reception during transmit because there is no oscillator input to the receive mixer. Conversely, the Clk2 output is switched off during receive.

A feature of the Quadrature Sampling Detector is that either the RF input, or the LO input, must provide two paths in 90-degree quadrature. This is normally applied at the Local Oscillator where it can be easily controlled for best performance. So, two oscillator signals are required, with the same frequencies but a precise 90-degree phase offset. Generating this quadrature Local Oscillator signal is always difficult. Analogue phase shift circuits have limited accuracy. Often a divide-by-4 circuit is used, to produce quadrature oscillator outputs from an oscillator input at 4x the reception frequency. This also creates challenges particularly as you try to increase the reception frequency to cover higher bands. For example, on 10m e.g. 30MHz, a local oscillator at 120MHz is required and the divide-by-4 circuit must be able to operate at such a high frequency. Devices such as the 74AC74 can do so, but pushing it higher into the 6m band cannot be done with the 74AC74.

The Si5351A has a phase offset feature, which is not really very clearly described in the SiLabs documentation. However, QRP Labs has perfected the technique to put two of the Si5351A outputs into precise 90-degree quadrature, which is maintained without tuning glitches as the frequency is altered. It's a nice development because it eliminates one more circuit block (the 74AC74 divide-by-4 circuit), again reducing complexity and cost. To the best of my knowledge this the first time the Si5351A has been implemented in a product directly driving a QSD with two outputs in quadrature (no divide-by-4 circuit).

5.4 Transmit/Receive switch

Since the receiver is entirely disabled during transmit, because of the absence of any local oscillator signals to the Quadrature Sampling Detector, the demands on the transmit/receive switch are considerably reduced. Now the circuit does not have to provide the massive amount of attenuation necessary to prevent the transmitter from overloading the receive circuits. All it has to do is provide a reasonable amount of attenuation, enough to stop the 5W signal (45V peak-peak) from damaging the receiver input mixer.

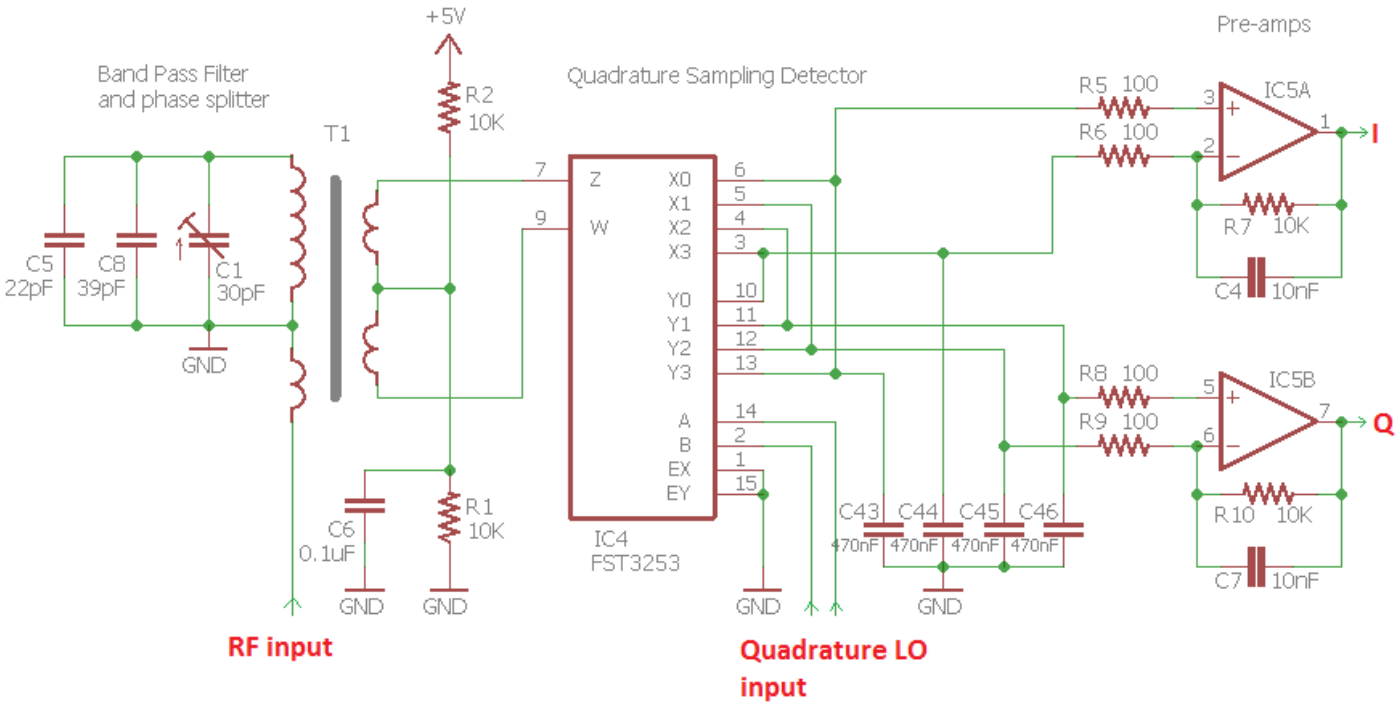


The transmit/receive switch is implemented by a single BS170 MOSFET. The source is at DC ground (via the primary of input transformer T1). The control signal from the microcontroller switches the MOSFET on or off. Interestingly, capacitor C34 close to the MOSFET gate is found to be necessary to prevent inductive pickup of the 5W RF from partially switching on the MOSFET.

The switch wouldn't provide enough attenuation to mute an operating receiver; but during receive, our receiver isn't operating; all the switch has to do is protect the Quadrature Sampling Detector from seeing 45V peak-peak which would destroy it.

5.5 Band Pass, Phase Splitter, QSD and pre-amps

Since the band-pass filter, Phase splitter, Quadrature Sampling Detector and pre-amp circuits are so tied up together, I am going to consider them all together in this section.



This circuit implements an input band-pass filter and double-balanced Quadrature Sampling Detector with low-noise pre-amps. Yet it does this with a low parts count, and resulting low complexity and cost. The FST3253 is a dual 1:4 multiplexer which is often seen in QSD circuits. It has fast switching times and very low on resistance of only a few ohms. The input signal is switched by the quadrature LO to each of the four integrating capacitors C43-C46 in turn, for 90-degrees of the RF cycle each. The result is that the audio difference (beat) between the RF input and LO input appears across each of the four integrating capacitors, with four phases at 0, 90, 180 and 270 degrees.

The operational amplifier IC5a takes the difference of the 0 and 180-degree outputs and amplifies it, resulting in the I output of the QSD. Similarly, IC5b differences the 90 and 270-degree outputs to produce the Q output.

The combination of the relatively large 470nF capacitors and the low source resistance results in a fast roll-off of the audio response. This is effectively a very narrow band pass filter since any incoming RF more than a few kHz away from the LO frequency is greatly attenuated. The QSD is therefore inherently a very high performance mixer design with high third order intercept and dynamic range, and low loss (0.9dB).

The FST3253 dual switch is often connected with the two switches simply paralleled together (which does half the switch ON resistance). But I prefer the double-balanced mixer configuration which provides higher performance. The double-balanced configuration requires two RF inputs 180-degrees out of phase (opposite to each other).

Despite the high IP3 and dynamic range, it is still prudent to provide some input band pass filtering to protect the mixer from strong out of band signals. In this CW transceiver design, the T1 transformer provides a simple solution to all of these problems with a very low parts count.

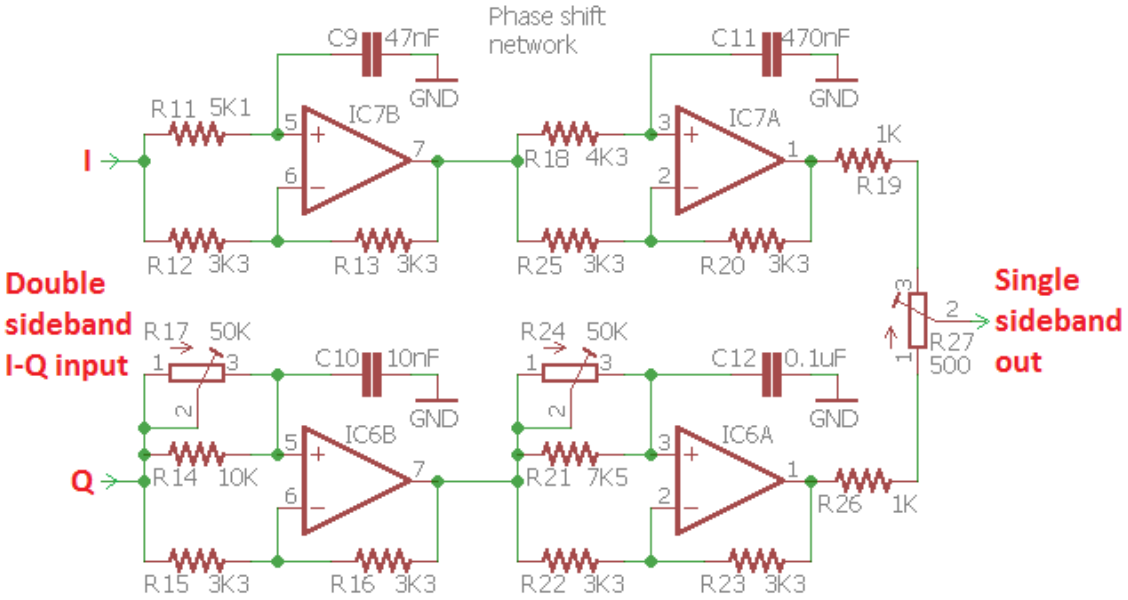
The primary couples the incoming RF into the two secondary windings which feed the double-balanced detector. One end of the primary is grounded which neatly keeps the DC potential of the input at ground, so the transmit/receive switch is easily implemented by a single MOSFET (see previous section). The two secondaries are connected as a centre-tapped single winding, which means that the outputs have 180-degree phase difference as required.

The secondary “centre-tap” is connected to a DC bias formed by R1, R2 and C6 at mid-rail i.e. 2.5V. This simple bias does not source or sink any significant current due to the balanced nature of the system, therefore no buffering is required. The DC bias feeds through the pre-amps, and into later stages – including the 90-degree phase shift network and the first three op-amps of the CW filter. It is a great benefit not to have to AC couple each stage with coupling capacitors, and then bias each stage individually. In this circuit, the same DC bias flows through all the way from this centre-tapped input transformer. This reduces component count and - you guessed it – complexity and cost. Another benefit may be that since much of the receiver signal path is DC coupled, it might be easier to reduce the inevitable “thumps” on switching between receive and transmit.

Finally, the band pass filter is implemented by a fourth winding on the same transformer T1, with some fixed capacitors and a trimmer capacitor forming the resonant circuit. It is only a single resonant circuit band pass filter so has limited stop-band attenuation, but it does have the benefit of low parts count, and simplicity of adjustment due to the single adjustment control.

5.6 90-degree audio phase shift

By this stage the I and Q outputs are each double-sideband, and we need to process them to demodulate single sideband.



The circuit used here is an active two-path all-pass phase shift network based on four operational amplifiers. The circuit is based on the same phase shift block as the Norcal NC2030 http://www.norcalqrp.org/nc2030.htm

In the real world, nothing is perfect – there are component tolerances to think about. The unwanted sideband suppression is maximised when the amplitude of the two paths is equal, and the 90-degree phase shift is accurate.

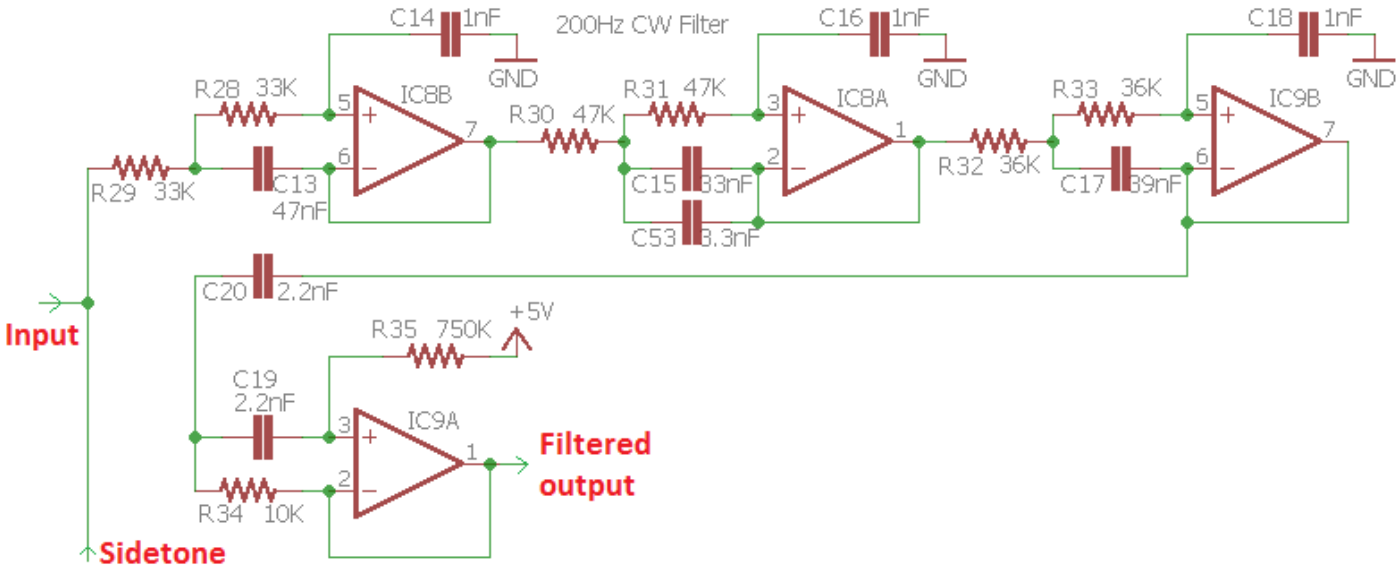
To improve the accuracy of the 90-degree phase shift, R17 and R24 allow adjustment of the phase shift at higher and lower audio frequencies respectively.

R27 allows adjustment of the balance between the I and Q channels, to equalise the amplitude from each path.

This CW transceiver kit includes built in alignment and test equipment, with a signal generator that can inject a test signal into the receiver input. It makes it easy to perform these adjustments, as described previously.

5.7 CW filter

The CW filter used in this receiver has a 200Hz bandwidth. The circuit is based on the HI-PER-MITE CW filter kit design by David Cripe NM0S, available from the Four State QRP group: http://www.4sqrp.com/HiPerMite.php (thanks David for permission to use it here). This is a high performance circuit specifically designed to avoid objectionable ringing.



There are three stages of low-pass filtering and one stage of high-pass filtering. The first three stages retain the 2.5V “midrail” bias all the way through from the input transformer T1. The final stage IC9A is biased using the 5V supply (avoiding a few extra components to create a real 6V midrail at half the supply).

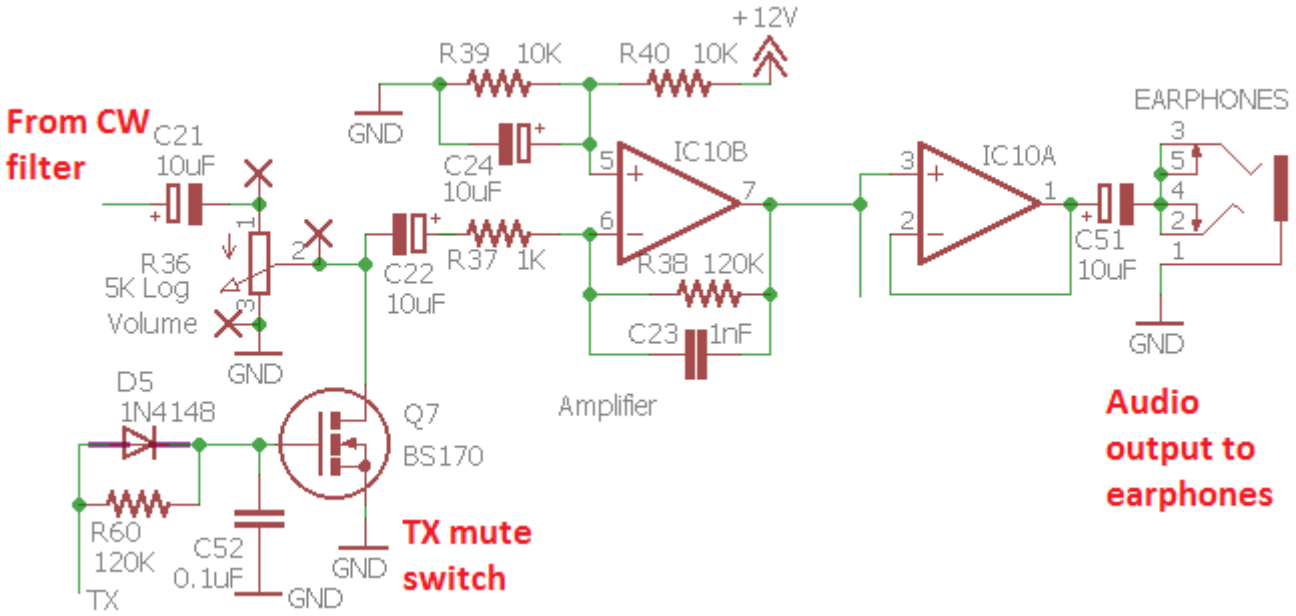
Sidetone is generated by the microcontroller and is a squarewave. To make it sound nice and clean, the sidetone injection is done at the input to the 200Hz CW filter, which cuts off any of the squarewave harmonics leaving a clean 700Hz sidetone.

The CW filter also provides a measured 18dB of gain.

5.8 Audio amplifier

The final stage in the receiver signal path is the audio amplifier, to drive earphones at a comfortable listening level.

R36 is a logarithmic potentiometer used as the gain control. With the wiper fully clockwise the receiver is at full volume. As the potentiometer is turned anticlockwise it forms a potential divider which attenuates the audio signal from the CW filter output.



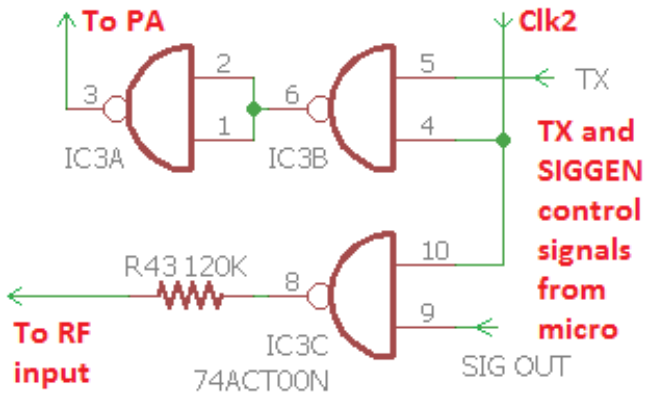
There is also a TX mute switch formed by Q7, another BS170 MOSFET. This was a late addition to the design: despite all attempts, I could not remove the nasty click on receive/transmit switching. The mute switch helps to attenuate it. The switch is operated by the microcontroller Receive/Transmit switch output. When the BS170 switch is on, it has a low resistance to ground which greatly attenuates the audio signal.

To reduce the audio “thump” when the transceiver is switched from transmit back into receive, the mute switch needs to remain switched on for a short while after the receiver is switched back on. A small wait while the thump subsides. This delay is achieved by the R-C network formed by R60 and C52. This would also introduce a delayed switch-on of the mute switch, which would allow the thump when switching to transmit to be heard. To prevent this, diode D5 was added, which bypasses the resistor R60 at the receive-to-transmit switchover. It ensures that at the receive-to-transmit event, the mute switch is enabled instantly; but on the transmit-to-receive switchover there is a short delay.

IC10B is a simple amplifier configured for 41dB of gain. The ½-V midrail bias is created by R39, R40 and C24. Using the 5V power line as “midrail” was found to add too much noise. Finally, IC10A is a simple unity-gain buffer. Although it is just an op-amp it is found to be perfectly adequate for driving standard earphones.

5.9 Transmit signal routing and PA driver

The 74ACT00 is a quad NAND logic gate. The input threshold voltage for a binary "1" is 2.4V which means that the gate is easily switched on by the ~3.3V peak-peak squarewave output from the Si5351A. The output of the 74ACT00 is 5V peak-peak, perfect for driving the BS170 MOSGETs in the Class-E PA into saturation.



The Clk2 signal from the Si5351A is used as the transmit oscillator as previously mentioned. It would have been easy to enable/disable the Clk2 output in software in the Si5351A chip configuration. However, this transceiver design also includes the built-in signal generator feature, for aligning the Band Pass Filter and adjusting the I-Q balance and phase adjustment controls. So, some of the spare gates in the 74ACT00 are used to switch the signal generator on/off and the RF Power Amplfier (PA) signal on/off, separately.

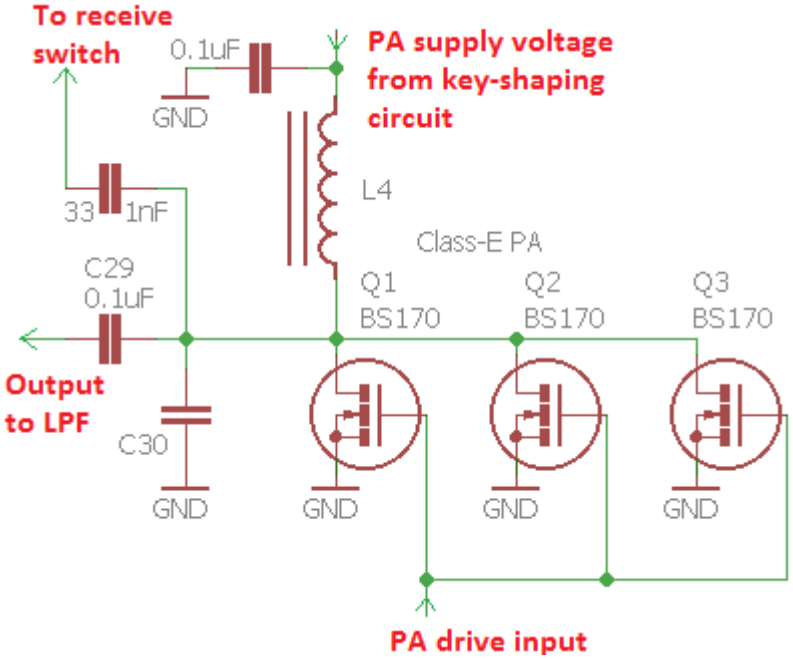
When the SIG OUT control line from the microcontroller is high, the Clk2 signal is enabled as signal generator, and routed via a 120K resistor straight to the RF input of the transceiver.

When the TX control line from the microcontroller is high, the Clk2 signal is routed to the PA. The final inverting gate IC3A is added to make sure that when the TX gate IC3B is off, the driver voltage presented to the BS170 gate is low, so the transistors are off.

5.10 Class-E Power Amplifier

A Class-E power amplifier is a wonderful thing. It has a very high efficiency, sometimes over 90%. This has several important benefits:

a) Since not much power is dissipated, we can use smaller (and cheaper) transistors
b) So little power is wasted as heat that the requirement for a heatsink is reduced or eliminated
c) During transmit the radio requires less current, so the drain on a battery is less – important for people who want to operate portable.



A Class-E Power Amplifier contains a resonant circuit at the frequency of operation, so it is only suitable for single-band use. A lot has been written about Class-E, much of it is very technical and mathematical.

Some excellent background reading are two papers by Paul Harden NA5N:

http://www.aoc.nrao.edu/~pharden/hobby/_ClassDEF1.pdf and
http://www.aoc.nrao.edu/~pharden/hobby/_ClassDEF2.pdf

Paul NA5N describes two defining features of Class-E:

1) Use of a square-wave drive to reduce switching losses: the transistors are either on, or off… no lossy region in between
2) Reducing the effects of the transistor capacitances. Class-E has a resonant tuned circuit. The capacitance of the transistors, normally an unpleasant lossy aspect, is now a part of the tuned circuit.

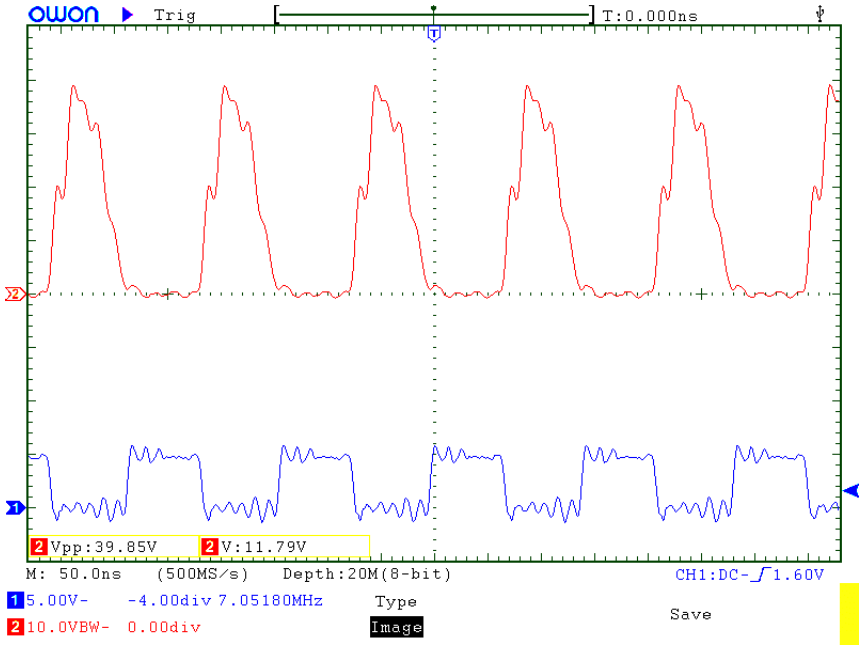
Class-E also has a reputation for being difficult to achieve. All those intense mathematics Google will help you find, don't help. In reality, once you realise the secret – it is not so difficult. Calculation of the impedance of a resonant circuit is simple, and there are many online calculators which will do the job for you. For example, http://toroids.info/T50-2.php which allows you to type in the operating frequency, and the desired resonant circuit impedance. Then the calculator computes the required inductance, capacitance, and the number of turns required for a certain toroid (in our case we use a T50-2).

The Class-E design process is simple. Choose the output impedance. We choose 50-ohms, because this is the input impedance of the Low Pass Filter we will use. The online calculator will tell you what inductance is needed, and how many turns to wind on the toroid. The online calculator also tells you the required capacitance to bring it to resonance at the operating frequency. Here we resort to experiment, because it is a little difficult to know what the output capacitance of the transistor is. The device capacitance varies depending on supply voltage and whether it is on or off. A simple experiment is required, adding different small capacitances to the circuit, and measuring the efficiency (measure supply voltage and supply current to calculate power input; then measure RF power output. Divide one by the other to get the efficiency). It is easy to find what additional capacitance is required to peak the efficiency. The resonance is quite broad and non-critical.

In this implementation, three BS170 transistors are used in parallel. The BS170 is inexpensive and small, but is rated for 500mA drain current and up to 830mW of dissipation. Per device. Three in parallel provides plenty of capability to achieve a 5W output on a single band.

There are always minor variations between device characteristics from one transistor to the next. If these were bipolar NPN transistors, we would not be able to parallel them in this way. If one transistor takes more of the load and starts to heat up, its resistance further decreases and this causes it to get even hotter. This process is known as "thermal runaway" and results (quickly) in destruction of the transistor. Emitter resistors are used to help balance the load. But with MOSFETs, their resistance INCREASES as the temperature goes up – so there is an inherent self-balancing when multiple devices are used in parallel, without any need for additional balancing resistors which would increase component count and waste some power.

This oscilloscope screenshot shows the classic Class-E waveform. Please ignore the ringing due to poor set-up of the 'scope probes etc. The lower (blue) trace is the 5V squarewave at the gate of the BS170 transistors. The upper (red) trace is the voltage as the BS170 drain. It peaks at approximately 40V in this example. This measurement was done with 12V supply and on 40m (7MHz).

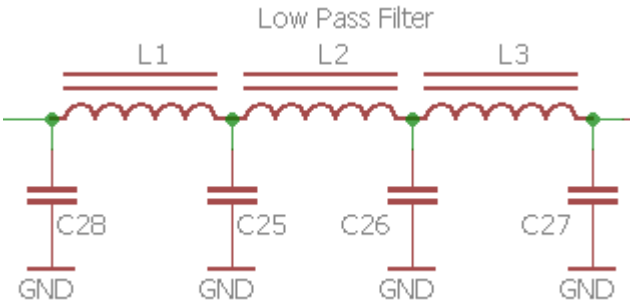


The important point to note is that when the BS170 are switched ON (the gate voltage is 5V), the drain voltage is zero. When the BS170 is OFF the gate voltage pulses nicely to a large amplitude. Class-E!

The summary: Class-E is actually quite easy to achieve in practice! Perhaps all the complicated mathematics might help to squeeze out another % or two of efficiency. But for practical purposes, it's a wonderful building block to use in a single-band CW transceiver.

5.11 Low Pass Filter

The transmitter output is rich in harmonics and must be followed by a good Low Pass Filter, to attenuate the harmonics and satisfy regulatory compliance.



The standard, well-proven QRP Labs Low Pass Filter kit http://qrp-labs.com/lpfkit is used here. To save space and cost, the components are installed directly on the PCB, not on a plug-in board.

It is a 7-element filter design originally by Ed W3NQN then published for many years on the G-QRP Club web site's technical pages.

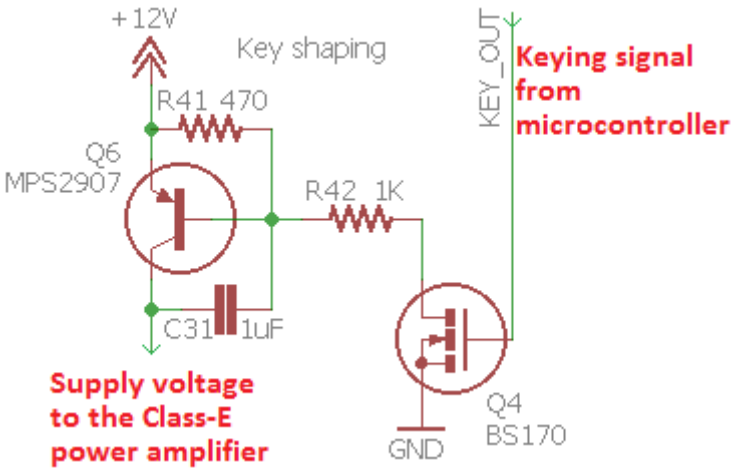
5.12 Key-shaping circuit

A hard-keyed CW transmitter generates clicks many hundreds of Hz away from the transmitted signal that can annoy users of adjacent frequencies. This is purely a consequence of the mathematics of the Fourier transform and is unavoidable. Any time you switch a signal instantly on or off, you WILL splatter energy onto unwanted nearby frequencies.

To combat this, any good CW transmitter should include an RF envelope shaping circuit to soften the key-down and key-up transitions. The ideal envelope shape is a raised cosine, but this is difficult to implement without significantly increasing the complexity of the circuit.

The simple key-shaping circuit used here uses only a few components but produces good results.

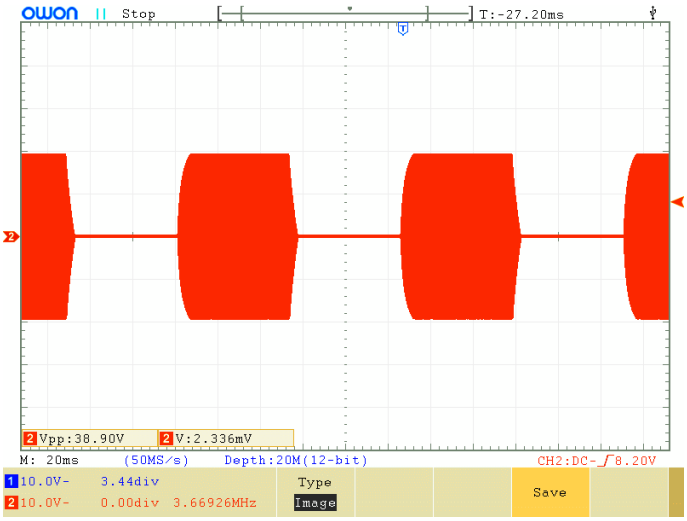
This circuit was derived by one published by Don Huff W6JL, see https://www.qrz.com/db/W6JL/ though as he says, "*this integrator-type keying circuit is found in many published homebrew designs over the past 40 years or so, so it is nothing new*". It uses a PNP transistor (Q6) and R-C integrator circuit. Don W6JL uses this key-shaping circuit to drive a 600W Power Amplifier.

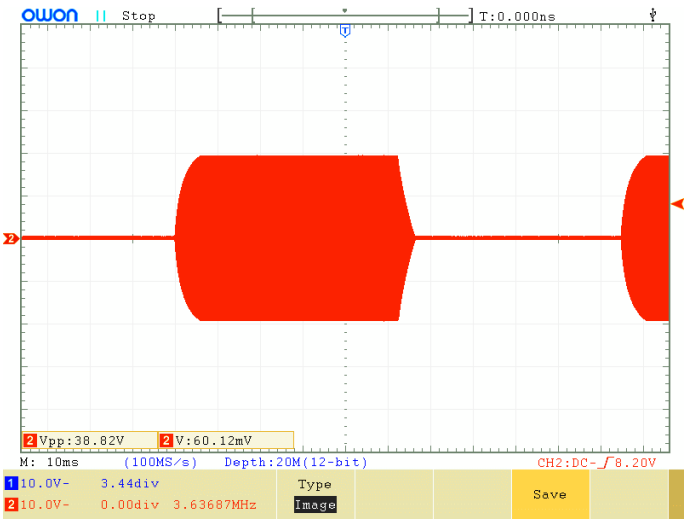


On key down the Q4 switch is "closed" by a high signal coming from the microcontroller. In a really simple transmitter, Q4 could just be replaced by a straight Morse key to ground! But in our case, the microcontroller implements automated stored message sending, beacon modes, and Iambic keyer – so we need the microcontroller to be the boss of everything. The microcontroller reads the state of the straight key or paddle, and processes it to produce a key output. When in straight key mode the microcontroller transfers the signal straight through from the key input, to the key output control line – but in other modes the processor must generate the keying signal.

The component values set the rise and fall time. With the components shown, the rise and fall time is about 5 milliseconds.

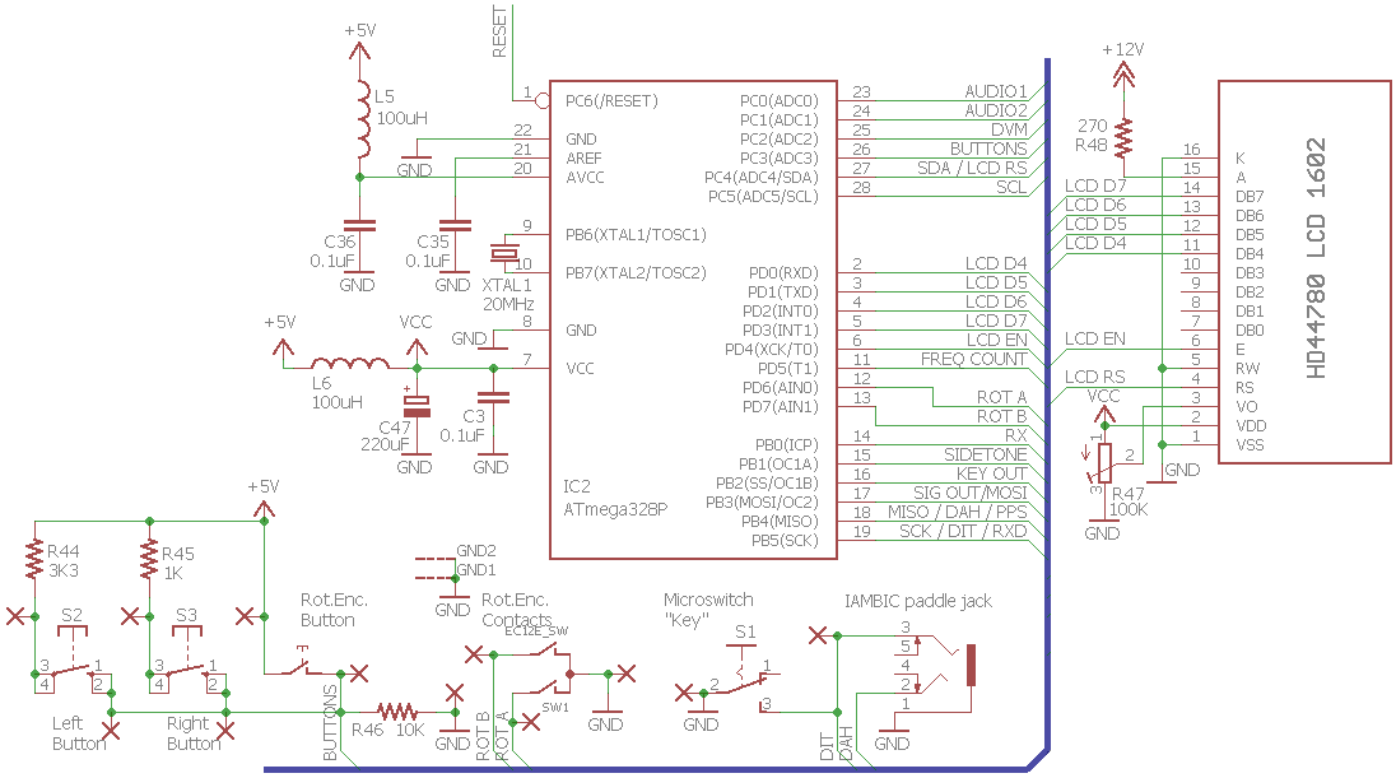
The following oscilloscope screenshots show a 40m band (7MHz) transmission, keyed with a continuous series of CW dits at approximately 24 words per minute. The amplitude is approximately 3.8W into a 50-ohm dummy load (with 12V power supply).





5.13 Microcontroller

The ATmega328 microcontroller circuit controls many aspects of this transceiver. Below is this section of the transceiver circuit. Several points are worthy of discussion.



ATmega328P processor

The ATmega328P was chosen because it has enough processing power and I/O to handle all the tasks required here. It is also common and inexpensive, and lots of QRP Labs products already used it, bringing economies of scale in both the kit preparation and the coding. The processor is operated at its maximum rated 20MHz system clock speed.

The code is all written in C and is not open source. While the same ATmega328 processor is used in the popular Arduino Uno products, there is no relation between code written for the Arduino environment and the custom code written for this CW transceiver.

Tuning clicks

Some constructors of radio receiver projects that use the Si5351A report loud clicks every time the frequency is changed. The cause of these clicks is one or both of two underlying issues:

a) Faults in the software configuring the Si5351A
b) Power line or radiated noise from the microcontroller/LCD back into the sensitive receiver

The first of these is not an issue here since we have already extensive experience using the Si5351A and have perfected its configuration.

The second issue is important to address. Every time the microcontroller updates the Si5351A configuration to cause it to change frequency, it typically also writes the new frequency to the LCD. There is a burst of activity in the microcontroller, and on the digital control signals to both the Si5351A and the LCD. The LCD controller chip will also be doing some work to effect the changed display. All of these digital transitions can radiate noise into the receiver front end. Changes in power consumption cause noise on the supply lines which can also be converted into noise detected in the receiver front end.

To combat the radiation issue, the ATmega328P microcontroller is sited at the back of the board, right under the LCD module. The tracks between the LCD module and the microcontroller are therefore kept as short as possible, to minimise radiated noise.

To keep noise out of the supply, the 5V supply to the microcontroller and LCD module is filtered by 100uH inductor L6 and 220uF capacitor C47 (may be supplied as 470uF, even better!).

In combination, these measures ensure that there are no “clicks” in the audio when tuning the receiver; just a small “flutter” as I call it, which is a natural consequence of the sudden change in frequency (Fourier rules).

Liquid Crystal Display module

The transceiver uses an HD44780-compatible LCD Module with 16 characters by 2 rows. The LCD is operated in the 4-bit mode in order to minimise the I/O pins used. No data is read back from the LCD which means the Read/Write pin can be grounded. In total only 6 I/O pins are used for writing to the LCD.

The usual contrast adjustment trimmer potentiometer is R47 and must be set to obtain a readable display.

The LCD backlight consumes about 30 or 35mA of current. The backlight could be connected directly to the 5V supply but this would somewhat increase the power dissipation of the 7805 regulator. In order to avoid overheating the regulator, this backlight is powered instead directly from the +12V rail via R48, a 270-ohm resistor. This resistor value was chosen so that over a supply voltage range of 7 to 20V the backlight brightness still appears reasonable. If you want a reduced brightness display, you could increase the value of R48.

Sidetone

In this transceiver, the sidetone is generated by Pulse Width Modulation using the ATmega328’s Timer1 peripheral. The frequency is configurable in the software via the configuration menu, and so is the volume. In order to control the volume, the microcontroller adjusts the duty cycle from 50% (maximum volume) down to under 1% (for minimum volume).

The sidetone is injected into the receiver path via a 3.3K resistor at the input to the CW filter. The sidetone generated by the microcontroller is a squarewave, rich in harmonics. As the volume is reduced, the duty cycle percentage drops and the amplitude of the 700Hz fundamental tone also drops. There are many harmonics of course, and the CW filter does a great job of removing these, so what is left in the earphones is a pleasant and pure 700Hz sinewave. This is why the sidetone is injected at the CW filter INPUT.

During transmit, when the sidetone is operational, the mute switch Q7 is also closed – but there is enough leakage through the imperfections of this switch that the sidetone gets through anyway. The 3.3K sidetone feed resistor R59 is chosen to pump enough signal through that it overcomes the attenuation of the mute switch. Without the mute switch, R59 would be a much higher value.

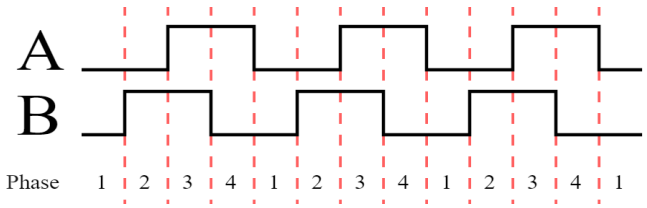
Key paddle, rotary encoder switch and buttons

The microcontroller keeps an eye on all the buttons, key paddle inputs, and rotary encoder switches. When button or switch closures occur as the operator activates a control, the microprocessor responds immediately as required.

The onboard microswitch “key” is wired in parallel with one of the paddle inputs. The paddle inputs, and the rotary encoder switch, are read using dedicated microcontroller I/O signals.

All mechanical switches exhibit switch bounce, where the switch contacts generate multiple transitions for a short time when the switch is activated. It is common to see in many projects, resistor-capacitor networks to debounce switches (including the rotary switch). Simple debounce circuits involving a resistor and a capacitor inevitably involve a compromise when choosing the R-C time constant. It is easy to miscalculate and make the time constant too short (bounce noise still gets through) or too long (rapid switch closures are missed). In some cases it is impossible to find the sweet spot in between these two extremes.

In my opinion, resistor/capacitor debouncing is a poor solution to the problem, when the circuit contains a microcontroller. It is easy to debounce the switch edges in software! This allows you to control time-constants or other debounce logic much more precisely. Of course, it saves those additional resistors and capacitors too, which all helps reduce the cost and complexity of the kit! This is the reason why in this kit, you will not see any resistor or capacitor debouncing of any of the controls. Everything is done in the software.



The rotary encoder is debounced using a state machine, rather than any timing loops. The rotary encoder produces two switch-closure outputs which have a 90-degree (quadrature, again) phase shift. The microcontroller can monitor these transitions and determine the number of “clicks” of the rotary encoder, and the direction the knob was turned in, and at the same time debounce the switch transitions.

Another unnecessary component often seen is a pullup-resistor on a switch signal to a microcontroller, so that when the switch is open (not pressed) the microcontroller sees a “high” voltage. This is unnecessary because modern microcontrollers such as the ATmega328P all contain configurable internal pull-ups!

I had a little more trouble when it came to the Left, Centre (shaft of the rotary encoder) and Right push-buttons. Despite thinking that the microcontroller had plenty of I/O signals available, when I came to implement all the features I had in mind, I realised there weren’t enough I/O signals to read the buttons!

The solution was to analogue-multiplex all three buttons onto a single input signal to the microcontroller, which is one of the Analogue to Digital Converter (ADC) channels. No microcontroller internal pull-up is used because I need to have precise control over the voltage levels. The input signal is pulled low by a 10K resistor R46. One side of all three buttons is connected to the I/O pin. The other sides of the buttons are connected to 5V via different resistors: 3.3K (R44) for the left button, 1K (R45) for the right button, and direct connection for the centre button. When one of the buttons is pressed, the resistors form a potential divider that sets the voltage to the microcontroller depending on which button is pressed. Some calculation results in the following table of voltages depending on which button is pressed:

Button pressed	**Voltage**
None	0.00V
Left	3.76V
Centre	5.00V
Right	4.55V

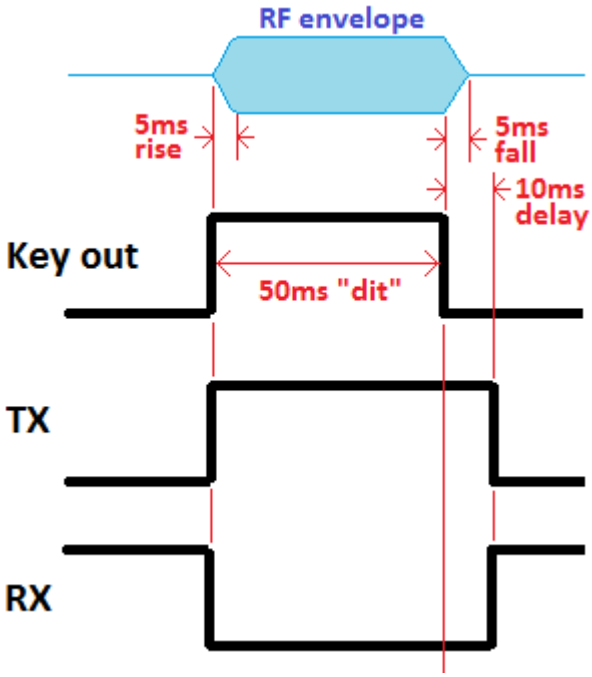
In all three cases, the voltage when a button is pressed, is higher than the 3V logic “1” threshold of the microcontroller. I am able to use a pin-change interrupt to detect that a button has been pressed, then read the ADC channel 3 and compare the measured voltage to the table, to determine which button was pressed.

It’s a nice trick for reading multiple buttons on a single ADC input, and it would work also for a larger number of buttons! Each one just needs a different resistor value and eventually, some care would need to be taken to ensure that the voltage differences could be read reliably, not masked by ADC noise or component inaccuracies. For three buttons, none of these problems arise.

TX/RX Switch control, and Key Out signals

As already discussed, the Key Out signal either reflects directly the state of a straight key, in Straight keyer mode, or it is a processed keyer signal when using the Iambic keyer modes or Beacon function of the transceiver. The Key Out signal is generated by the microcontroller with precise timing in the Beacon and Iambic keyer modes. It is routed to the RF envelope key-shaping circuit discussed previously.

The Transmit/Receive switch must remain in “transmit” state for a short time after key-up, in order for the RF envelope of the key-shaping circuit to drop gently down to zero, which takes approximately 5ms. Due to this, the microcontroller implements a 10ms delay between key-up and switching the radio back to receive mode.



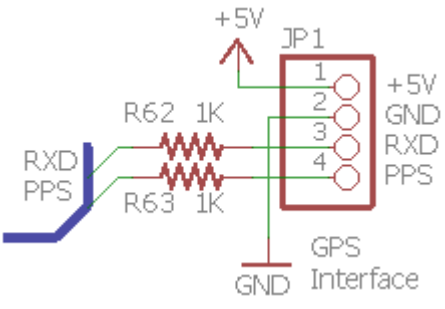
The diagram (right) illustrates the relationship between these signals, for an example 24 words per minute “dit”. The Key out signal has a “high” period of 50ms exactly. The ramp-up and ramp-down of the RF envelope has 5 millisecond (approximately) rise and fall times. The “RX” control line generated by the microcontroller switches on the receiver (allows RF through to the receive mixer) when it is “high”. It is held in the “low” (transmit) state for 10ms after key up completes.

The “TX” signal is the inverse of the “RX” signal, and it switches the gate of the audio mute switch Q7 during transmit. One of the NAND gates IC3D is used as the inverter to create the “TX” signal from the microcontroller’s “RX” output.

5.14 Optional GPS interface

The GPS was a late addition to the design. It is added to facilitate several useful features:

- Calibration of the 27MHz reference oscillator
- Calibration of the 20MHz system clock
- For WSPR, discipline of the frequency (drift correction)
- For WSPR, setting and discipline of the real time clock
- For WSPR, setting the Maidenhead Locator square



The optional GPS interface is provided by a 4-pin header at the bottom right of the board, next to the 3.5mm socket for the keyer. The pinout of the 4-pin header exactly matches the one on the QRP Labs Ultimate3S QRSS/WSPR transmitter kit, and the QRP Labs QLG1 GPS receiver kit. This makes connecting the radio to the QLG1 GPS kit very easy, if you wish. However other GPS modules could also be used. The following requirements of the GPS module must be met:

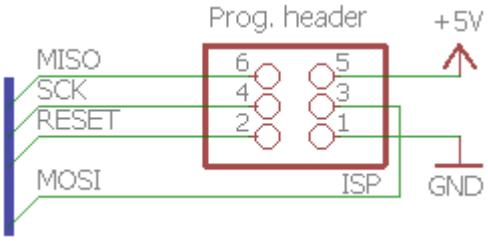
- Positive-going (leading edge) 1 pulse per second signal with a reasonable pulse width – e.g. some older GPS module have a 1us pulse and this will not be long enough. The QLG1 GPS receiver provides a 100ms (0.1s) duration pulse, though shorter pulses would also work.
- 9,600 baud serial data sending standard NMEA 0183 GPS sentences
- "TTL" voltage levels i.e. 0V "low" and 5V "high" (or at least, below 1.5V and above 3.5V)

The 1pps and Serial data signals from the GPS are connected via 1K resistors, to the "dit" and "dah" inputs to the microcontroller. The GPS is therefore wired directly in parallel with the paddle. The GPS may not be used at the same time as the paddle. This would cause the transmitter to be keyed by the 1pps and serial data signals. The 1K resistors protect the GPS outputs in case the paddles or onboard microswitch "key" inadvertently ground them.

The ATmega328P has a USART serial data peripheral. However, the pin assignments were already made, so far down the design process, at the time the GPS receiver option was included. Furthermore, the I/O pin availability was already scarce. So, the two GPS signals share the same I/O as the paddle inputs. A software-emulated USART peripheral module now reads the GPS serial data, at 9600 baud.

5.15 In Circuit Programming (ISP) interface

A 2 x 3-pin header is installed on the PCB to facilitate In Circuit Programming (ISP) of the ATmega328P microcontroller in future, if new firmware versions become available.



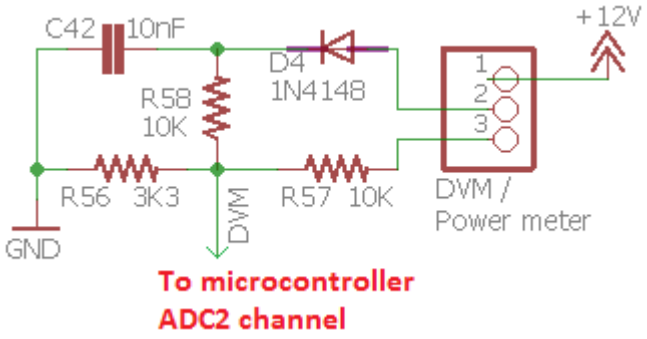
Firmware updates can be done with a simple USBISP AVR programmer available from eBay for under $2 including international shipping. The 2 x 3-pin header pinout matches the standard AVR programmer cables.

The ISP interface uses the same I/O pins as the GPS interface and the Paddle. When upgrading the firmware, be sure to disconnect the GPS (if you had one connected). And do not press any paddle key while doing the firmware update!

5.16 Test Equipment

The built-in test and alignment equipment is a really nice feature of this CW transceiver kit.

DVM and RF Power meter



The DVM and RF Power meter is implemented using the 10-bit ADC channel 2 input to the microcontroller.

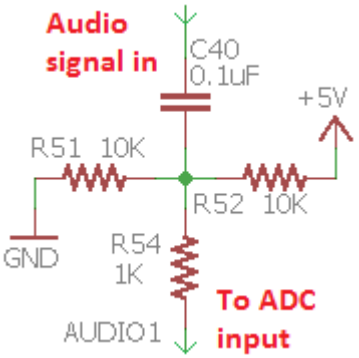
When measuring voltage, connect the voltage to be measured to pin 3 of the 3-pin DVM/RF Power connector. The voltage is dropped by the potential divider formed by R56 (3.3K) and R57 (10K). This applies a division by a factor of 4.03. Since the ADC can measure voltages up to 5V, this means the DVM function has a range of 0V to 20V maximum, and each ADC bit has a resolution of 0.02V.

When measuring RF power, connect the RF to be measured to pin 2 of the 3-pin DVM/RF Power connector. In this case D4 and C42 form a simple RF detector. The RF is rectified by D4 and smoothed by C42. R56 (3.3K) and R58 (10K) again form a potential divider that divides the rectified voltage by a factor of 4.03. The resulting range of the power meter is from 0 to 5W approximately.

For normal operation of the radio, if you wish to see a battery voltage icon on screen, then connect pin 1 to pin 3 of the 3-pin header. This connects the voltmeter input to the 12V supply of the radio.

In the unlikely event of accidentally overloading the RF power input or the DVM input, the 10K resistors R57 and R58 limit the damage to the ATmega328P processor or maybe prevent damage altogether. Perhaps. (don't feel tempted to try it intentionally).

Audio ADC

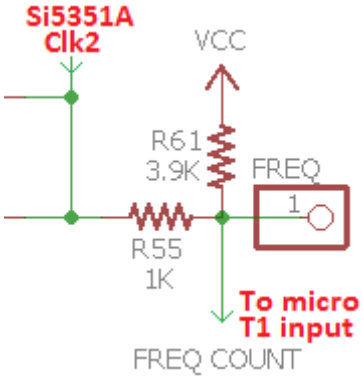


There are two audio connections from the receiver signal chain to the microcontroller. One is connected at the output of the I-Q balance adjustment potentiometer. In other words, the input to the CW filter. The second audio connection is connected after the final audio amplification gain stage. Both of these audio signal channels are fed to Analogue to Digital Converter (ADC) inputs of the microcontroller.

The circuit diagram (right) shows the circuit at AUDIO 1 (ADC channel 0) which is connected to the audio amplifier output. A duplicate circuit is used on the other channel. The purpose of the circuit is to remove the DC bias from the audio signal, and apply a new bias of 2.5V, to bias it to the middle of the ADC's input range. The 1K resistor in series with the ADC pin is intended to provide some kind of limited protection to the microcontroller in the event that the voltage falls outside the range 0-5V.

The ADCs in the ATmega328P have 10-bit resolution which translates to about 5mV. The ADC conversion result is a number in the range 0-1023, with 512 representing zero. An audio voltage of +/- 2.5V can therefore be measured. In practice, the audio signal is much less than this, which means the full ADC range is not really used. The AUDIO2 channel (connected at the CW filter input) has even less amplitude so is not used for anything except the BPF Alignment procedure, which has a very strong signal.

Frequency Counter



The frequency counter test pin input is connected directly to the microcontroller's 16-bit Timer1 input pin.

There is a 1K series resistor (R55) to connect it also to the Si5351A's Clk2 output. This is used during the optional GPS-disciplined 27MHz reference crystal frequency measurement. In this function, the Si5351A is configured to route its raw 27MHz output, divided by 4, to its Clk2 output. This passes through the 1K resistor R55, to the microcontroller. The 1K resistor provides protection to the Si5351A in the event you accidentally have something connected to the FREQ input pin, and the Si5351A is operating its Clk2 output at the same time.

Since the 3.3V peak-peak signal from the Si5351A is rather marginal for operating the Timer1 input, a 3.9K pullup resistor increases its centre point. Now "low" is 1.02V and "high" is 3.65V, which is enough to trigger the Timer1 input reliably.

The frequency counter can operate from 0 to 8MHz theoretically, assuming a clean 50% duty-cycle squarewave signal of sufficient amplitude. The 8MHz upper bound is a limitation of the AVR processor which has a synchronous timer input, and can only count to 40% of the system clock frequency (40% of 20MHz is 8MHz).

The signal to be counted must satisfy the logic thresholds of the ATmega328P, namely a "low" or "0" is less than 1.5V, and a "high" or "1" is greater than 3.5V.

Signal Generator

The signal generator mode was also discussed in the operating manual section. The signal output is simply taken from the Clk0 or Clk1 output of the Si5351A. Care should be exercised not to draw excessive load (e.g. short circuit) from these pins, which could damage the Si5351A. It will be very difficult to replace a damaged Si5351A.

Self-test of Signal Generator and Frequency Counter

Note that once the Signal Generator has been activated by scrolling to the Signal Generator menu item, and pressing the left (select) button to enable it – the Signal Generator remains active on the set frequency, until you press the right (exit) button and leave the Test Equipment menu.

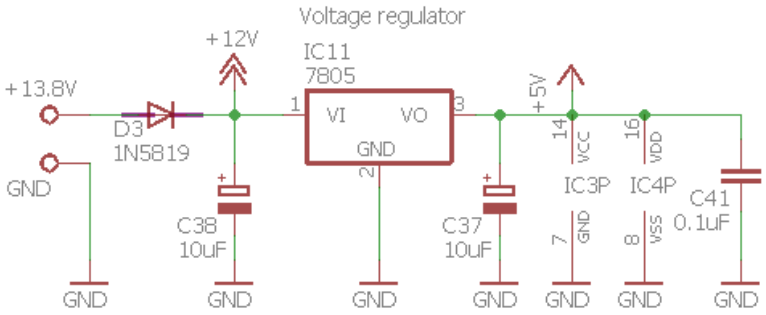
Press the left button (select) to enable the signal generator; an underline cursor appears under the digit that will be tuned by the rotary encoder. Adjust the signal generator to produce a frequency some way under 8MHz. Let's say, 6MHz. When you leave edit mode, by pressing the left (select) button until the cursor disappears off the right side of the 1's digit, OR by pressing the right button (exit), then the cursor will no longer be shown on the display. Turning the rotary encoder 1 click anticlockwise shows menu 9.5, the frequency measurement. Now if you connect a wire from the FREQ input to the CLK0 output, you will measure the signal generator's frequency. This is shown here in the photograph (right); clearly the 2kHz low reading indicates I have not calibrated the 27MHz reference value and/or the 20MHz system clock value!

In the photograph, the white wire shown looping around the rotary encoder connects the FREQ pin (just under the “M” character of the LCD, in the photograph) to the Clk0 output pin (just under the “z” character of the LCD).

This is a convenient way to get the frequency counter to verify the correct operation of the signal generator (and hence the Si5351A synthesiser). The signal generator output is available at both the Clk0 and Clk1 pins, hence you can test them both.

5.17 5V voltage regulator

5V for the microcontroller, Quadrature Sampling detector chip, LCD module and IC3 (quad NAND gate) is provided by IC11, a 7805 5V 1A voltage regulator.

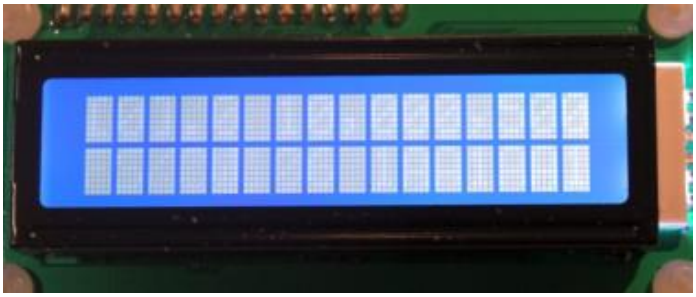


Note that the input is protected against reverse polarity by D3, a Schottky diode type 1N5819. This also drops the voltage by 0.3 to 0.4V, which has the effect of reducing the output power.

If you are brave, and confident that you will NEVER connect power to the radio with reversed polarity, you can omit this diode and fit a jumper wire over it.

6. Fault-finding

6.1 Blank LCD or blocked LCD



If your LCD screen looks like either of these, then it is a sure sign that you have not adjusted the contrast trimmer potentiometer R47.

6.2 No backlight at all

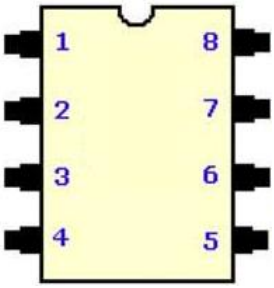
If there is no LCD backlight at all, check that you have installed R48, and also check that you have properly installed the LCD connector between the main PCB and the LCD module.

6.3 A row of blocks appears on the top row

If you see a row of blocks along the top of the LCD module, and the bottom row is empty, this means that the microcontroller has not communicated with the LCD module.

Check that your IC2 microcontroller is correctly installed in its 28-pin socket, with the correct orientation (dimple on the chip matches the dimple on the socket and the PCB silkscreen). Check that all of the pins of IC2 are correctly inserted in the socket, not bent.

6.4 DC voltage readings



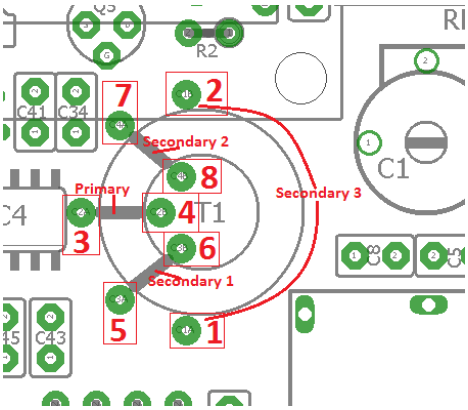
The following table lists the DC voltage read at various points around the circuit. This may aid troubleshooting. The voltage measurements were made using the in-built DVM, in menu item 9.1. These measurements were made using a 12V power supply.

Where an IC pin number is mentioned, remember that pin numbers start from the top left (just to the left of the dimple at the end of the chip) and are counted anticlockwise from there (see diagram of 8-pin IC for example, right).

Where “left”, “right”, “top” and “bottom” are mentioned, they assume the PCB is orientated in the normal way.

All measurements were made probing the PCB from the top side. In some cases, poke the wire to touch the pad of a particular capacitor or resistor.

References to the pinout of transformer T1 use the nomenclature indicated in the diagram (right).



In many cases these measurements may not exactly match your own. In particular if you use a power supply other than 12V, then voltages which are derived from the 12V supply line will be different. For example, bias voltages on op-amps which are biased to half the supply rail.

Bear in mind that in some of these measurements, the act of measurement will itself change the measurement. The “input resistance” of this DVM is only 13.3K (R56 and R57 in series, assuming high input resistance of the microcontroller pin). If you were to make these DC measurements with a high input resistance “real” DVM rather than the inbuilt DVM of this radio transceiver, these measurements would change. Again: if your measurement is in the right ballpark, do not worry if it differs from these.

When making these measurements, be very careful that whatever wire you are using to probe the circuit does not cause shorts to adjacent pins or components, which could cause damage.

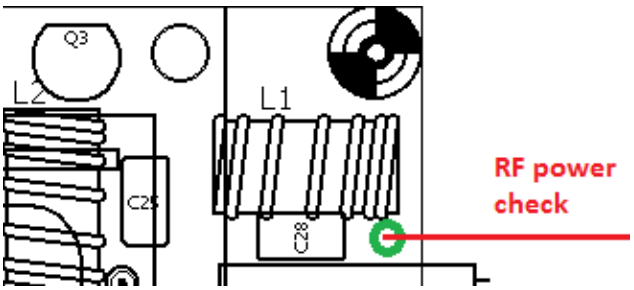
I suggest avoiding directly probing the Si5351A chip pins, they are just too small: the risk of shorting adjacent pins is high. You can easily probe nearby components which are connected to the Si5351A pins, if you wish – refer to the PCB tracks diagram. Similarly, I do not recommend directly probing the FST3253 pins. Again, it is easy to probe nearby components which are connected to those pins, and reduces the risk of damaging something by causing inadvertent short circuits.

Location	Voltage	Comment
+ supply terminal	11.98	As I already mentioned: a 12V supply
Clk0 test pin	~1.70	The Clk0 pin in normal operation has a 3.3V peak-peak squarewave with 50% duty cycle. The DVM is measuring the average of that. Measuring here makes a lot of nasty noise in the audio.
Clk1 test pin	~1.70	The Clk1 pin in normal operation has a 3.3V peak-peak squarewave with 50% duty cycle. The DVM is measuring the average of that. Measuring here makes a lot of nasty noise in the audio.
Left side of C2	3.40	This is the voltage of the Si5351A supply, pins 1 & 7
Right side of D2	4.96	This is the +5V regulated supply
T1 pins 1, 2, 3 & 4	0.00	
T1 pins 5, 6, 7 & 8	2.36	
Top side of C43	2.28	C43-46 are the QSD integrating capacitors. The voltage across each one differs slightly. The following measurements show these small differences.
Top side of C44	2.44	
Top side of C45	2.42	
Top side of C46	2.30	
IC5 pin 1	2.64	IC5a output, the pre-amp I output
IC5 pin 2	2.46	
IC5 pin 3	2.26	
IC5 pin 4	0.00	
IC5 pin 5	2.28	
IC5 pin 6	2.46	
IC5 pin 7	2.44	IC5b output, the pre-amp Q output
IC5 pin 8	11.67	This is the supply voltage minus the voltage drop caused by the reverse polarity protection diode D3
IC6 pin 1	2.44	IC6a output
IC6 pin 2	2.44	
IC6 pin 3	1.63	
IC6 pin 4	0.00	
IC6 pin 5	1.55	
IC6 pin 6	2.44	
IC6 pin 7	2.44	IC6b output
IC6 pin 8	11.67	
IC7 pin 1	2.64	IC7a output
IC7 pin 2	2.64	
IC7 pin 3	1.99	
IC7 pin 4	0.00	
IC7 pin 5	1.89	
IC7 pin 6	2.64	
IC7 pin 7	2.64	IC7b output
IC7 pin 8	11.67	
IC8 pin 1	2.46	IC8a output
IC8 pin 2	2.46	
IC8 pin 3	0.65	
IC8 pin 4	0.00	
IC8 pin 5	0.67	
IC8 pin 6	2.48	
IC8 pin 7	2.48	IC8b output

IC8 pin 8	11.67	
IC9 pin 1	4.84	IC9a output
IC9 pin 2	4.84	
IC9 pin 3	0.65	
IC9 pin 4	0.00	
IC9 pin 5	0.67	
IC9 pin 6	2.44	
IC9 pin 7	2.44	IC9b output
IC9 pin 8	11.67	
IC10 pin 1	5.83	IC10a output
IC10 pin 2	5.83	
IC10 pin 3	5.83	
IC10 pin 4	0.00	
IC10 pin 5	4.21	
IC10 pin 6	3.94	
IC10 pin 7	5.83	IC10b output
IC10 pin 8	11.67	
LCD pin 1 (leftmost)	0.00	LCD VSS supply pin
LCD pin 2	4.94	LCD VDD supply pin
LCD pin 3	0.57	LCD Constrast
LCD pin 4	4.92	LCD RS pin
LCD pin 5	0.00	LCD RW pin
LCD pin 6	0.00	LCD E pin
LCD pin 7	1.08	LCD DB0 pin
LCD pin 8	1.08	LCD DB1 pin
LCD pin 9	1.06	LCD DB2 pin
LCD pin 10	1.04	LCD DB3 pin
LCD pin 11	0.00	LCD DB4 pin
LCD pin 12	4.94	LCD DB5 pin
LCD pin 13	4.92	LCD DB6 pin
LCD pin 14	0.00	LCD DB7 pin
LCD pin 15	4.41	LCD backlight anode
LCD pin 16 (rightmost)	0.00	LCD backlight cathode

6.5 RF Power output check

If you go into menu “9.2 RF Power”, you can check your RF power output. Your RF output should preferably be connected to a dummy load, since the RF power calculation assumes that the voltage is across a 50-ohm load.

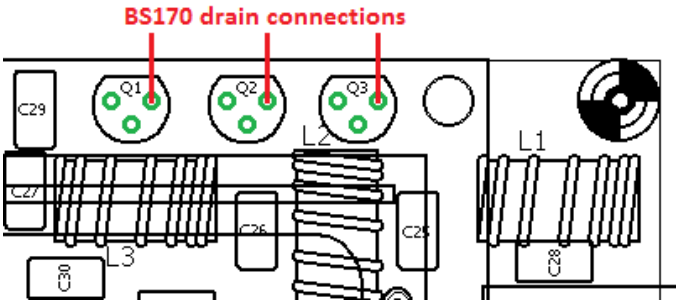


At the top right corner of the PCB, you will find a pad on the PCB where the L1 wire is soldered in. This pad is connected directly to the RF output pin of the BNC connector (see diagram). If you connect a wire from here, to pin 2 of the 3-pin DVM/RF Power header, then key-down, the display will read the power output in Watts. It is easiest to have the transceiver in Straight key mode and just press the microswitch.

Remember that the simple diode RF detector is not likely to be particularly accurate. But if you see a reading of a few watts, it will give you confidence that your transmitter is working properly.

If you do NOT see any power output on key-down, then you probably have a fault in the Power Amplifier. A common mistake, as I mentioned previously during the assembly steps, is failure to remove or burn off the enamel insulation on the enamelled copper wire wound onto the toroids.

Check the RF power wire touched to the BS170 drains first (see diagram). You can still touch the wire to these BS170 drains from the top side of the board. Key-down and check the wattage reading. At this point the indicated power will be even less reliable than across the dummy load. But at least see if it reads e watts.



If you see power output at the BS170 drains, but not at the overall transceiver RF output, then this means that the fault lies in the Low Pass Filter. Check the soldering and the enamel situation, on those toroids.

If you do not even see power output at the BS170 drains, then perhaps the L4 drain inductor in the PA has incorrectly soldered wires, that have not made an electrical connection – so check the L4 wires first and make sure the enamel is properly scraped or burnt off, to make a proper connection.

To probe other parts of the Power Amplifier circuit is less easy, because the LCD module prevents easy access. You will need to prop the PCB up vertically, and reach behind with the wire, while trying to still be able to watch the LCD screen from the front.

6.6 Audible continuous clicking noise during receive

This is not a fault of the radio, it is a consequence or limitation of the simple design. When the LCD is updated by the microcontroller, the activity on the signal lines to the LCD and in the microcontroller, causes digital noise which propagates into the sensitive radio receiver. In most cases when the display is updated, for example, when you change the frequency, this weak click is completely inaudible. In any case altering the frequency results in an inevitable audio discontinuity as the received spectrum suddenly shifts to the new LO frequency, and this more than masks any weak click from updating the display.

The continuous clicking noise during receive, sounding like the motor of a boat (perhaps), is due to the display update for the S-meter. You are likely to only notice this noise when the antenna is disconnected, or you have a dummy load connected. Ordinary band noise normally massively exceeds the weak clicking.

In some cases (depending on your power supply, antenna, grounding etc) the interference may be worse than normal. If this is the case, you can just switch off the S-meter. Remember that the S-meter indication is anyway only a rough indication, and with a little experience your own ears will do a much better job of estimating Signal level of the received station. The S-meter can be switched off in the “Show S-meter” configuration parameter in the “Other” menu.

7. Measurements

The measurements in this section were performed on assembled kits. There will always be differences from one kit to the next, due to component tolerances, adjustment differences, how you wound toroids, etc. In addition, it is often difficult to make good RF measurements. Therefore,

everything you read in this section should be interpreted as an INICATION only. Your actual results may (and probably will) differ. The following information just gives a guide of very roughly, what you might expect to see.

7.1 Equipment

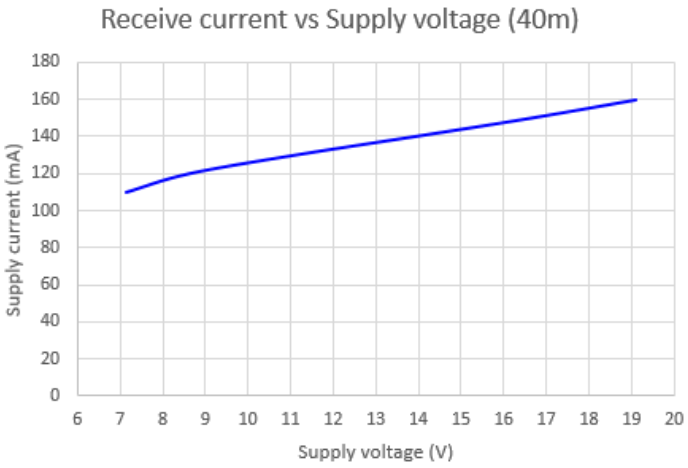
For reference, and for the sake full disclosure, the following equipment was used in making these measurements:

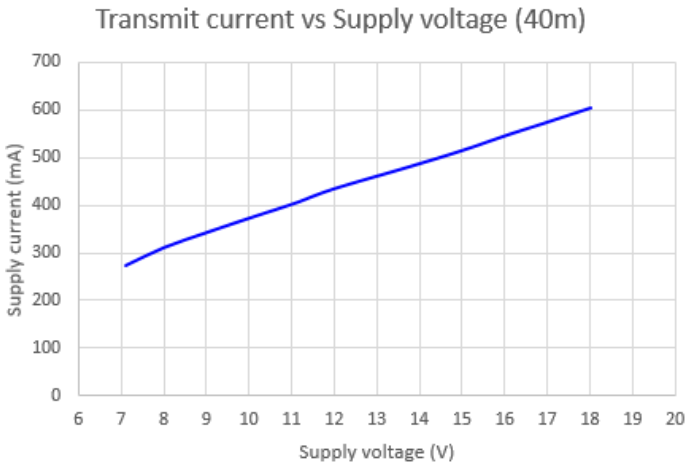
- OWON XS3102A 12-bit 100MHz Digital Storage Oscilloscope
- ADVANTEST R3361C 2.6GHz Spectrum Analyser with 50-ohm tracking generator
- QRP Labs 50-ohm Dummy Load kit http://qrp-labs.com/dummy
- XONAR U5 24-bit external USB sound card
- Argo and Spectran audio analysis software from http://weaksignals.com
- QRP Labs Ultimate3S for test signal generation http://qrp-labs.com/ultimate3/u3s
- Simple 14.000MHz battery-powered crystal oscillator signal generator
- Two generic yellow DVMs
- 12V fixed power supply, and variable voltage bench supply

7.2 Transceiver current consumption

The current consumption of your radio may be an important consideration if you intend portable battery-powered operation. Current consumption depends on supply voltage.

These measurements were carried out on a 40m band version of the kit. They will not differ much for other bands. At 13.8V supply for example, current consumption is about 140mA on receive, and almost 500mA on transmit.





At supply voltages above 13.8V, check the temperature of the 7805 voltage regulator, which may need a heatsink as it dissipates more heat at higher supply voltages.

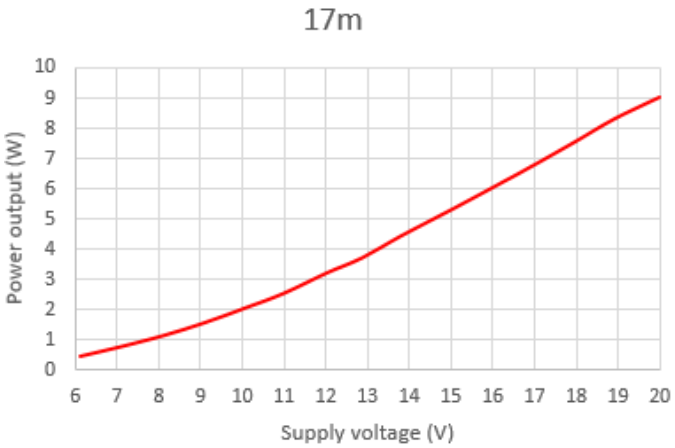
7.3 Transmitter power output

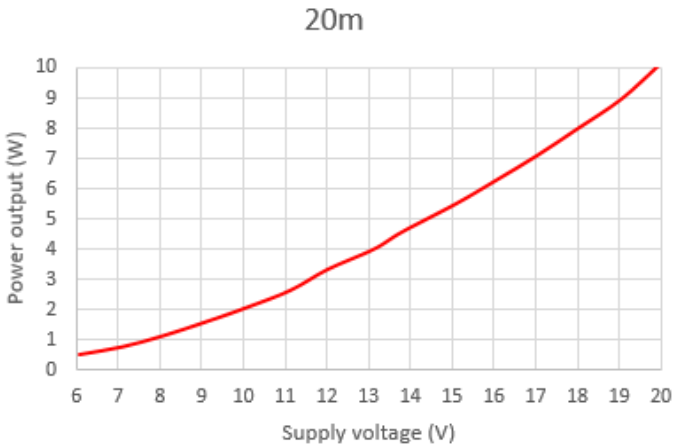
The transmitter RF power output varies depending on the power supply voltage. It is also dependent on band. Your results may vary depending on your Low Pass Filter inductor construction! If the cut-off frequency becomes too low, then you can start to get attenuation at the operating frequency. In this case you can remove a turn or two from each toroid.

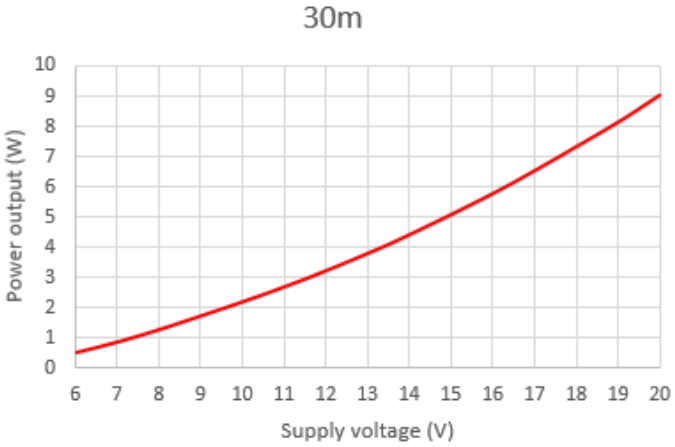
The following charts show the power measurement using the oscilloscope to measure the peak-peak amplitude across the 50-ohm dummy load.

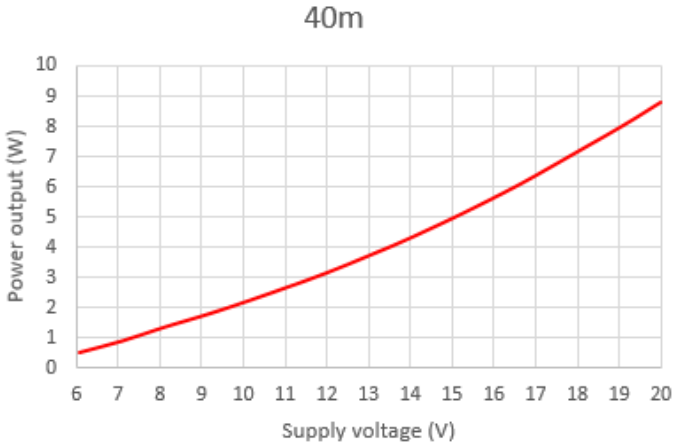
Operation at output power levels above 5W is NOT recommended, as it will put too much strain on the BS170 output transistors which may result in failure. For this reason, limiting the supply voltage to not more than 15 or 16V is recommended.

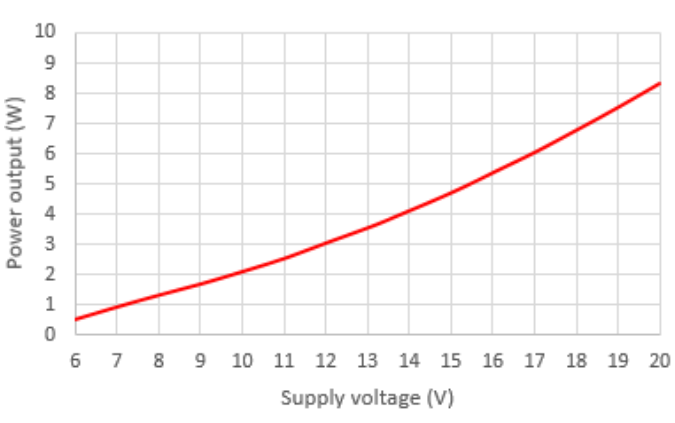
Note that these measurements do not take into account the D3 1N5819 reverse polarity protection diode voltage drop (approximately 0.3 or 0.4V). In other words, the measurements were done in a transceiver with D3 bypassed by a wire link.

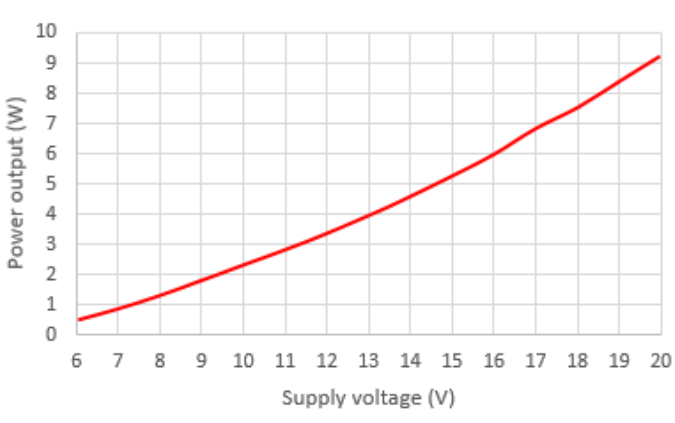








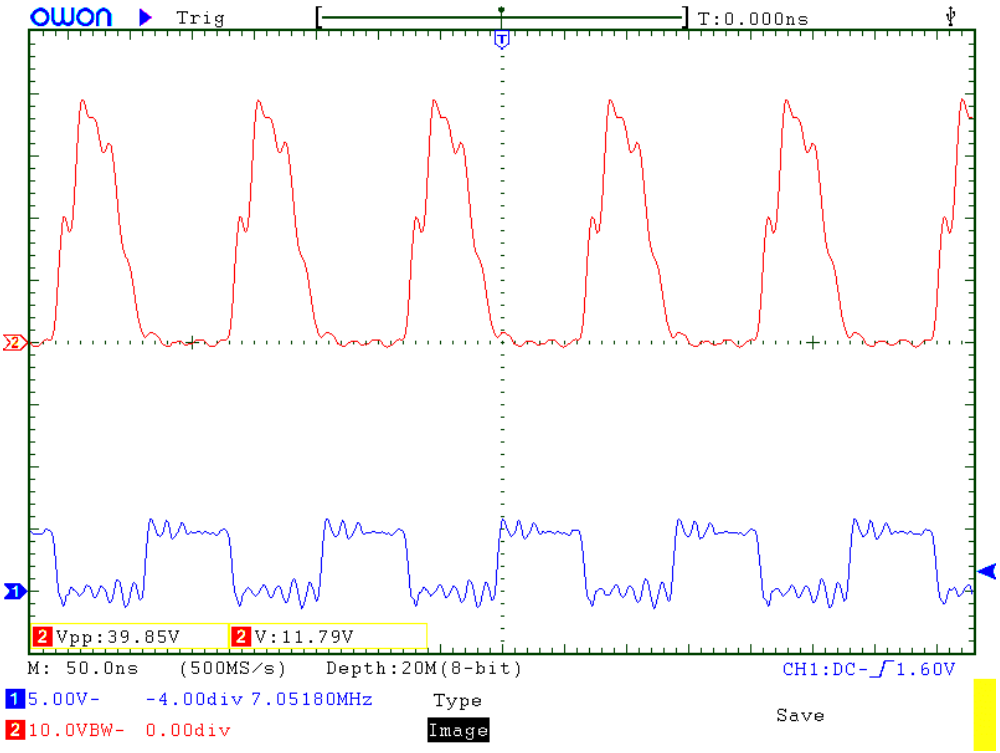




7.4 Class-E Power Amplifier drain waveform

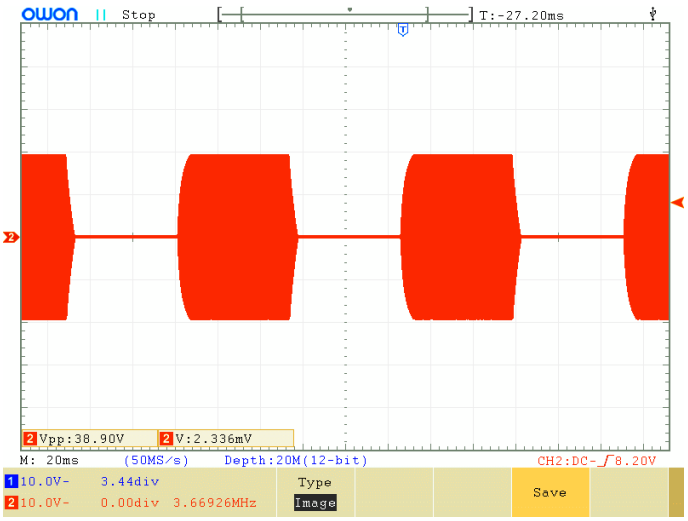
This oscilloscope chart shows the waveform at the BS170 drains (top, RED colour trace) and the input drive waveform, a 5V peak-peak squarewave (bottom, BLUE trace). The 40m band is shown.

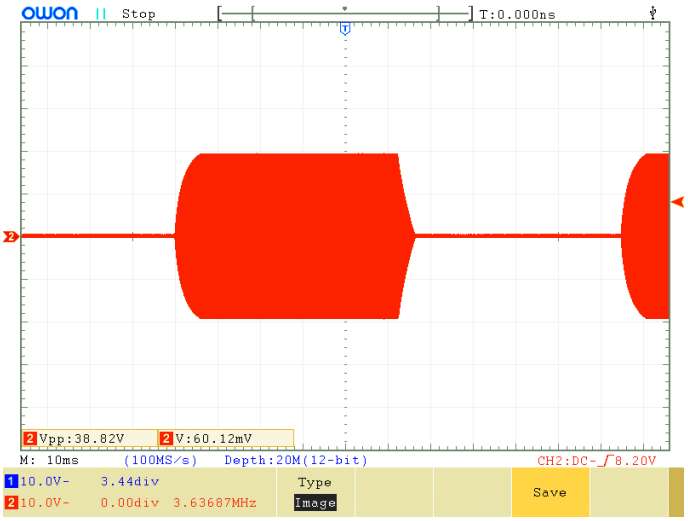
Upon ignoring the “ringing” artefacts due to poor ‘scope probes etc., the waveforms are correct for Class-E operation.



7.5 RF envelope key-shaping

The following oscilloscope images show the RF envelope when keyed with a continuous 24 words per minute (50ms duration) string of dits. The simple key-shaping circuit described earlier results in rise/fall times of around 5ms and significantly attenuates key-clicks.





7.6 Low Pass transmitter harmonic output filter characteristics

The Low Pass Filter response was not measured. The Low Pass Filter is the same as the QRP Labs Low Pass Filter kit module http://qrp-labs.com/lpfkit . Measurements of the Ultimate3S kit http://qrp-labs.com/ultimate3/u3s using these filters is on this page http://www.qrp-labs.com/ultimate3/u3info/u3spec.html

7.7 Band Pass receiver input filter characteristics

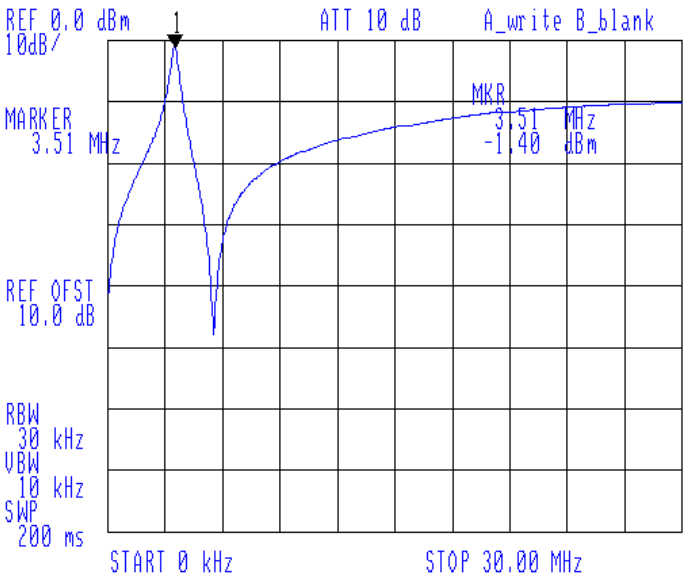
The following charts show measurements of the simple band pass receiver input filter implemented using transformer T1. For each band, one image shows the close-in response, the other shows the response across 0 to 30MHz.

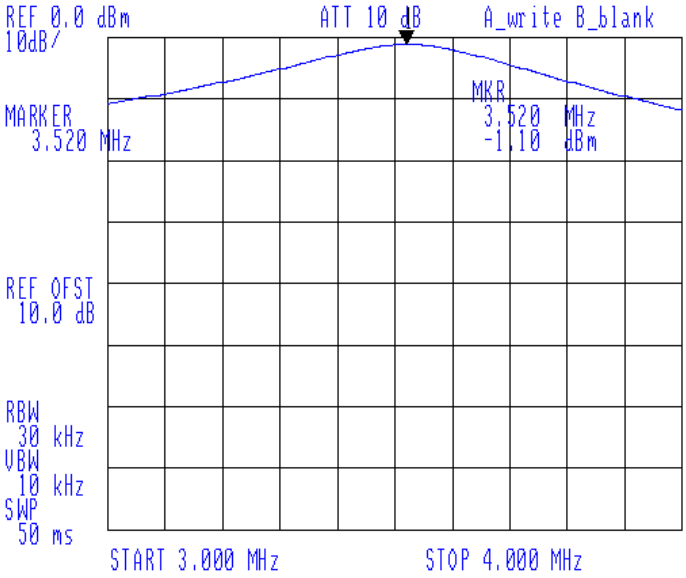
Note that in the circuit, the Band Pass filter sits behind the Low Pass filter. Therefore, the low pass filter response is ADDED to the band pass filter response. These charts show a relatively poor attenuation above the passband – but this is not a problem because in this region the Low Pass Filter provides high attenuation.

For each band, the 3dB bandwidth was measured. The tuning range available was also measured (from having the C1 trimmer capacitor plates fully meshed, to having them fully un-meshed). Also shown is the attenuation at the centre frequency.

80m band

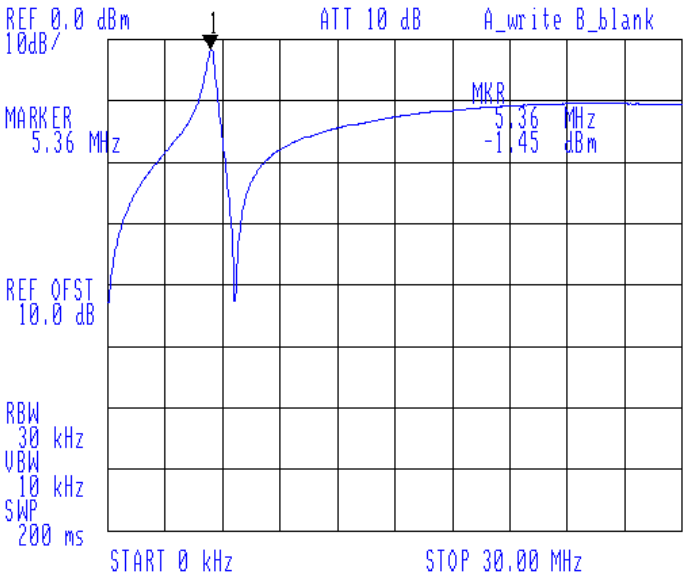
3dB bandwidth: 343kHz
Insertion loss: 1.10dB at 3.520MHz centre
Tuning range: 3.320MHz to 4.03MHz

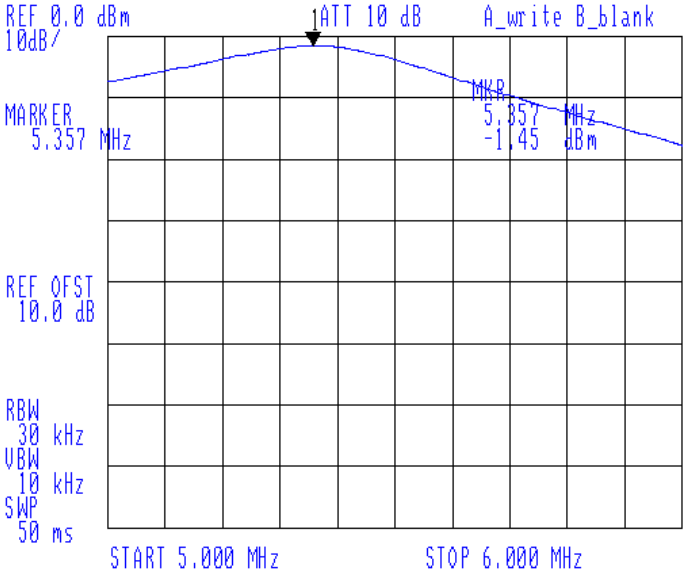




60m band

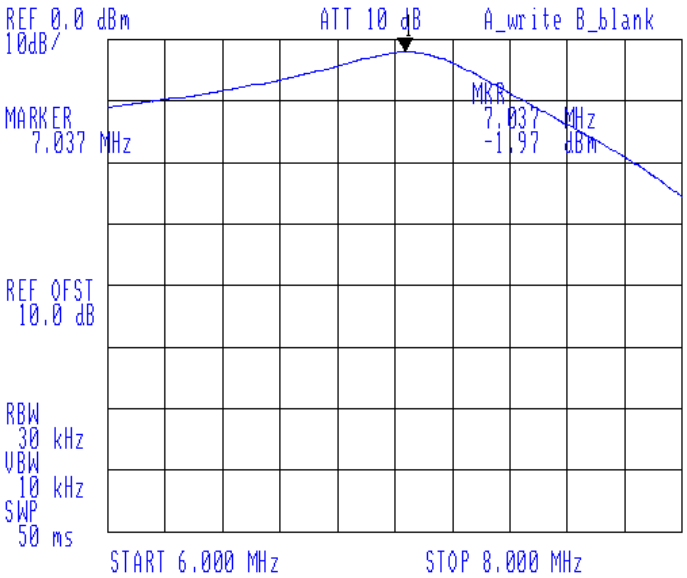
3dB bandwidth: 367kHz
Insertion loss: 1.45dB at 5.357MHz centre
Tuning range: 4.63MHz to 5.91MHz

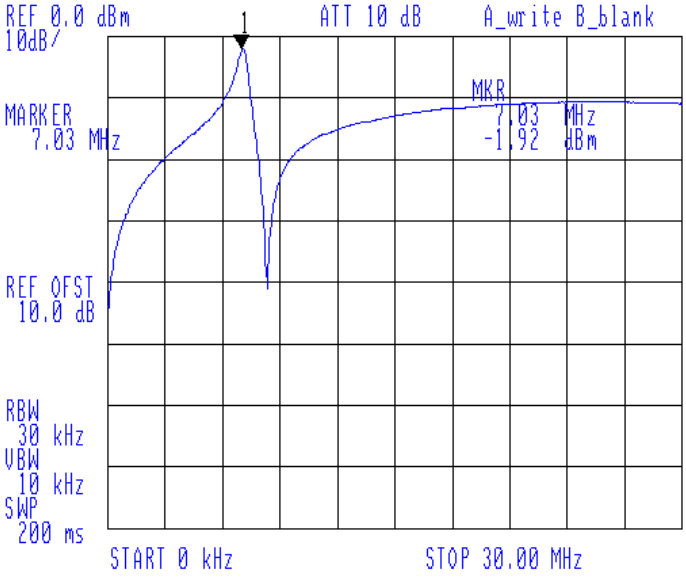




40m band

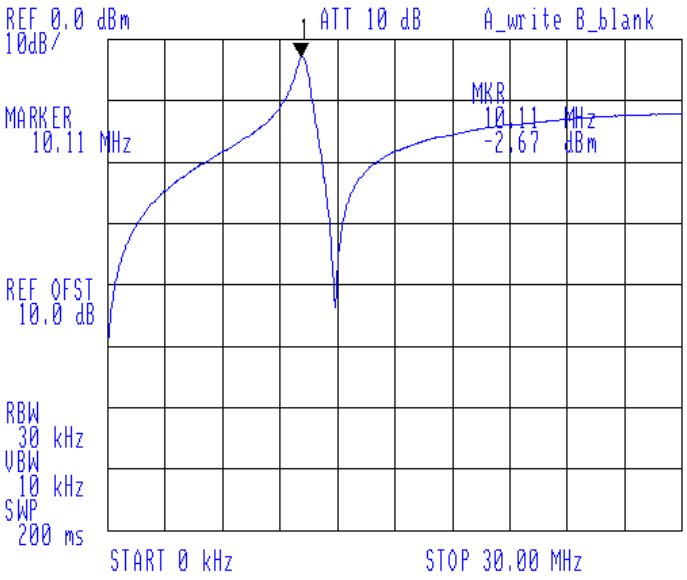
3dB bandwidth: 514kHz
Insertion loss: 1.97dB at 7.020MHz centre
Tuning range: 6.04MHz to 8.36MHz

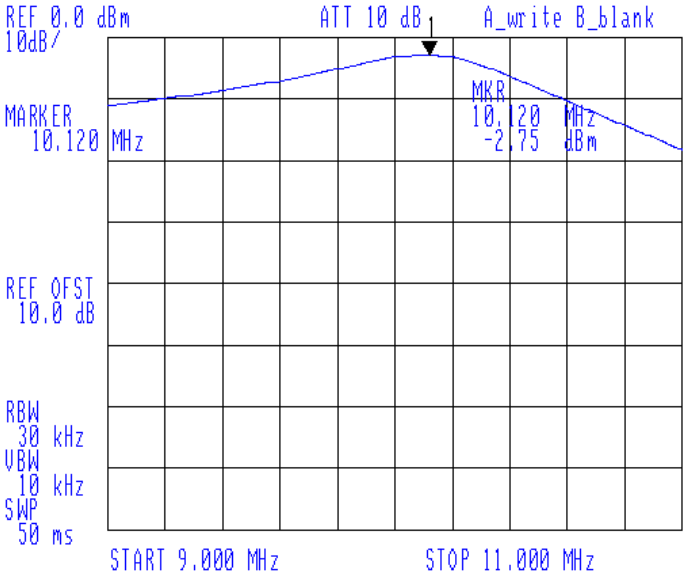




30m band

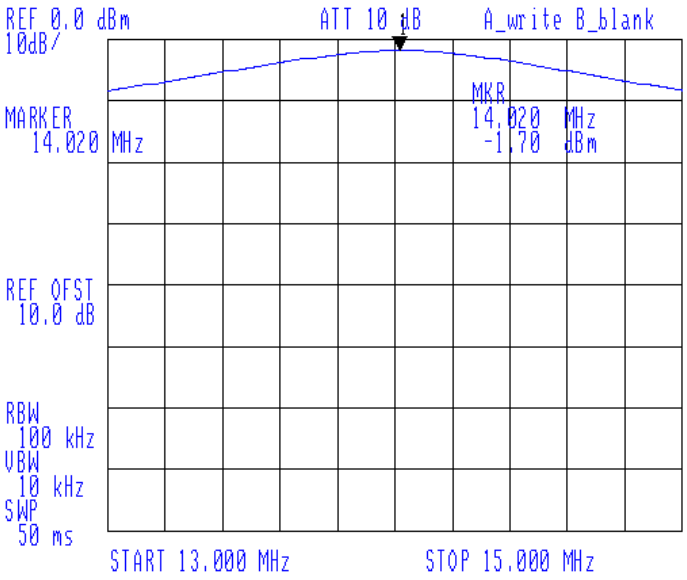
3dB bandwidth: 645kHz
Insertion loss: 2.67dB at 10.120MHz centre
Tuning range: 7.93MHz to 12.47MHz

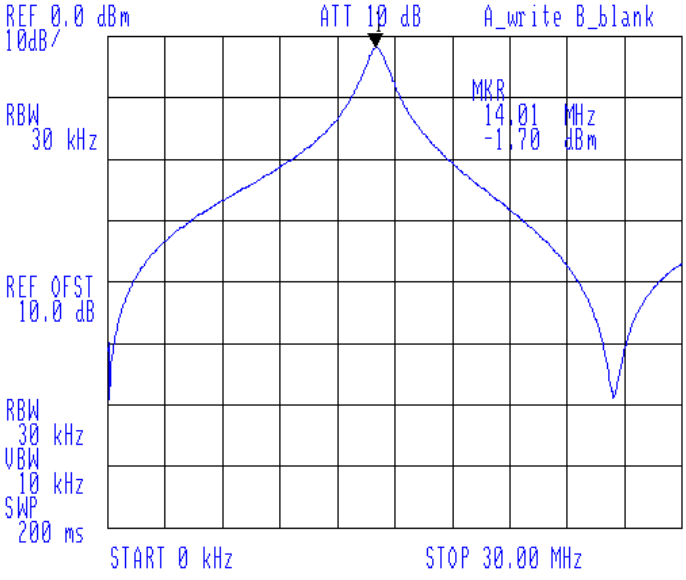




20m band

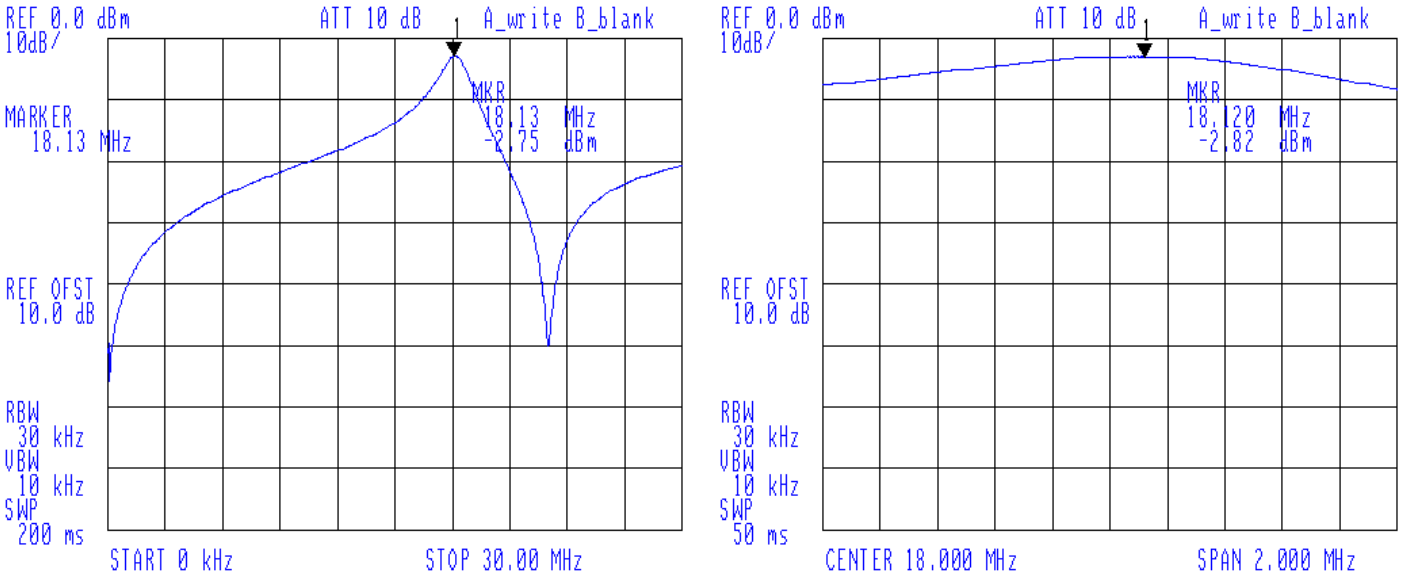
3dB bandwidth: 1,083kHz
Insertion loss: 1.30dB at 14.020MHz centre
Tuning range: 9.69MHz to 22.54MHz





17m band

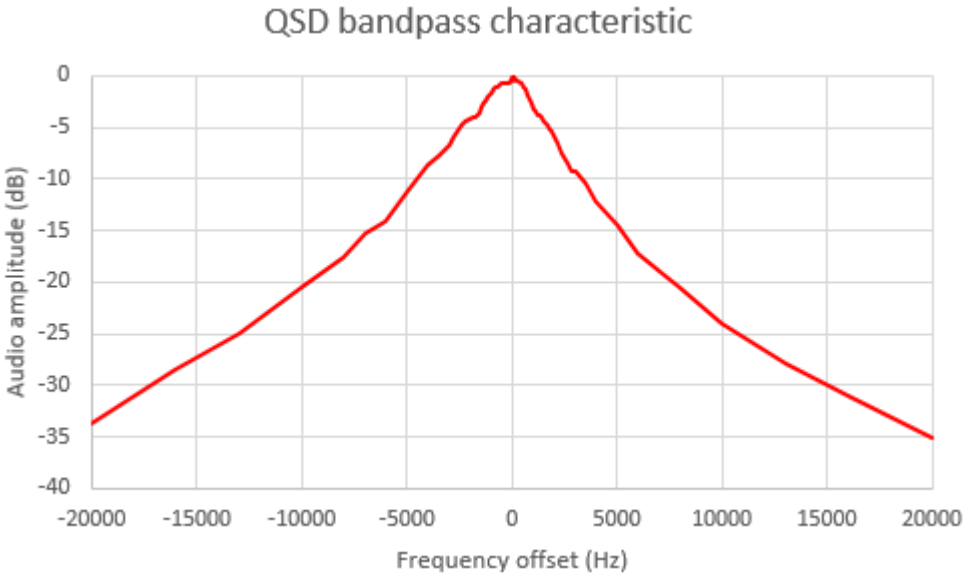
3dB bandwidth: 1,352kHz
Insertion loss: 2.77dB at 18.120MHz centre
Tuning range: 12.64MHz to 35.0MHz



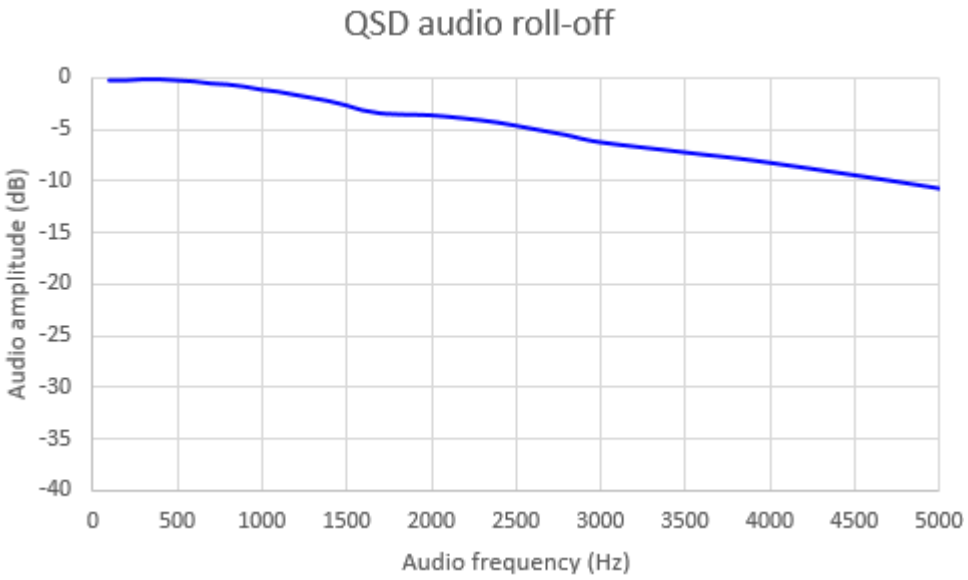
7.8 Quadrature Sampling Detector bandwidth

The following chart shows the attenuation naturally provided by the Quadrature Sampling Detector from -20kHz to +20kHz from the Local Oscillator frequency. The narrow characteristic of this circuit is an advantage because it effectively adds an additional narrow RF bandpass filter to the receiver, preventing strong nearby signals from reaching the audio amplifier stages. This improves intermodulation performance.

The circuit parameters are optimised for CW operation in this transceiver. A constructor wishing to use the I-Q outputs from IC5 to feed a computer SDR, would wish to flatten this response. To do this, reduce the values of the four 470nF capacitors C43 to C46.

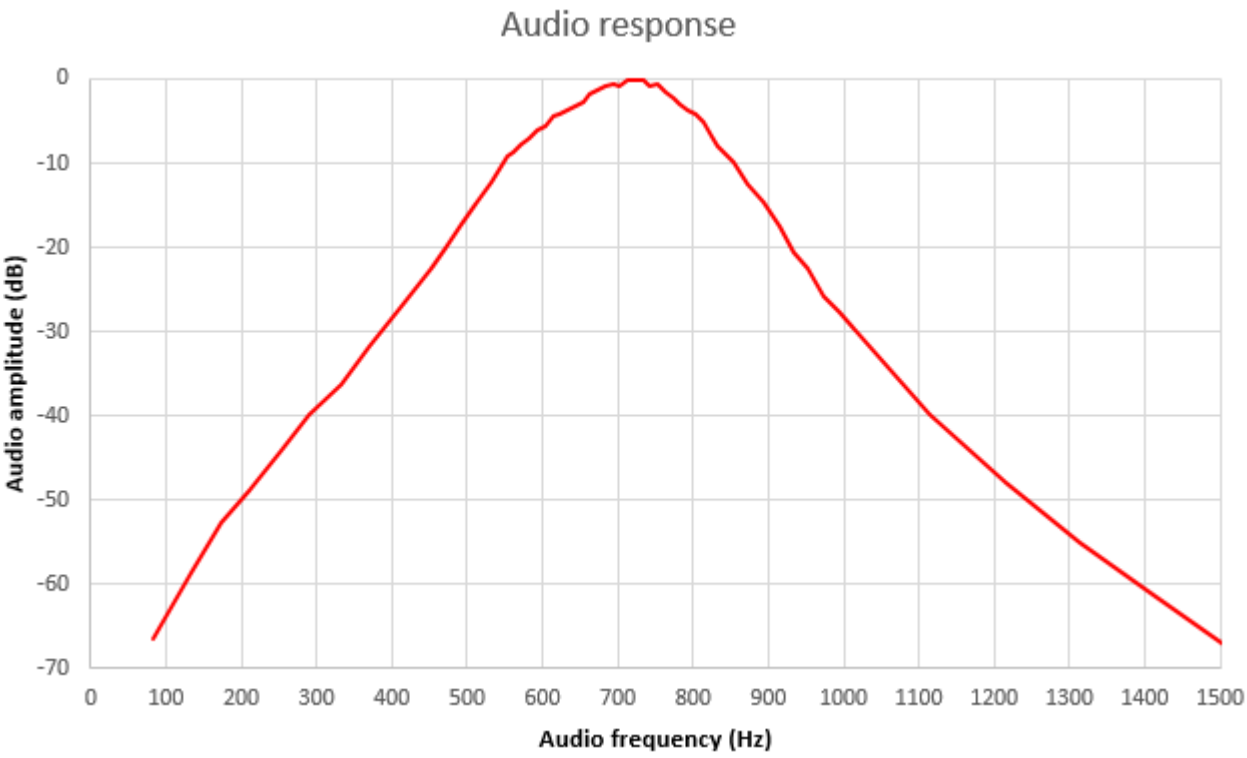


The following chart shows the Quadrature Sampling Detector roll-off over the narrower range 0 to 5kHz. You can see that at the CW operating frequency 700Hz, the attenuation is insignificant; the roll-off of the QSD does not significantly improve the selectivity of the CW filter, but it does improve the intermodulation characteristics of the receiver.



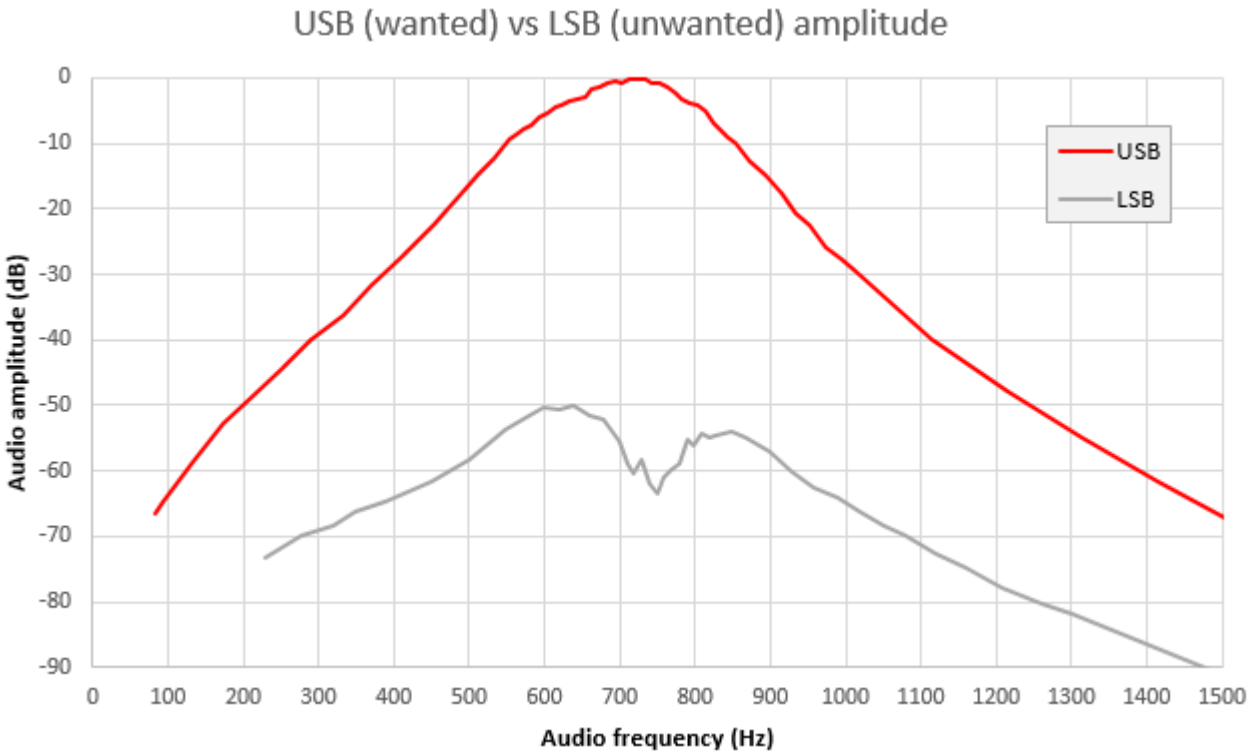
7.9 CW Filter response

This chart shows the measured response of the CW filter (combined with the other filtering in the receiver chain). It was measured by tuning the receiver across a test signal. Note that the centre frequency, a little over 700Hz, could be altered by changing some component values.

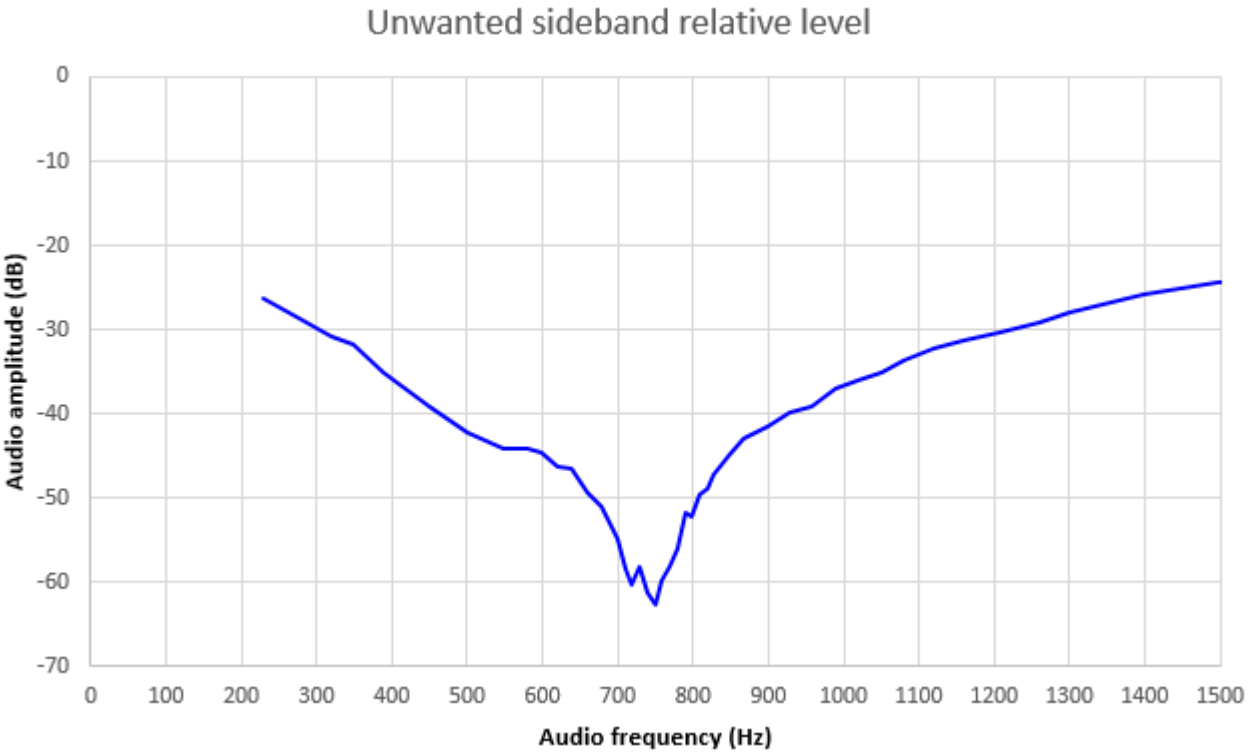


7.10 Unwanted sideband rejection

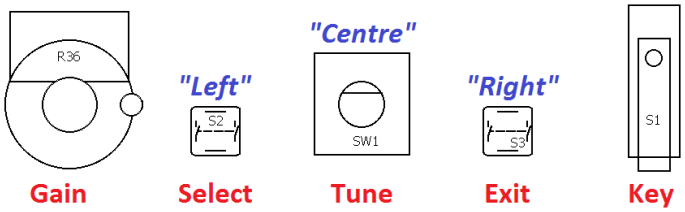
The following chart shows the measured level of the Upper Sideband signal (USB) and the unwanted Lower Sideband signal (LSB) when tuning the receiver through a strong test signal. The curves depend very heavily on the I-Q balance and audio phase shift adjustments, and these curves are from one measured prototype.



Subtraction of these two curves results in the unwanted sideband level as shown in the following chart. Normally anything above 40dB rejection is considered good; above 50dB is excellent. With the built-in adjustment features of this kit it is relatively easy to achieve really excellent unwanted sideband rejection.



8. Operation reference "cheat sheet"



Main controls functions:

Left single press: Keyer speed adjust, then left again to select, or right to cancel. **Left double press**: RIT adjust, then left again to select, or right to cancel **Left long press**: Enter the configuration menu (see below)	**Encoder turn**: tuning, menu selection, editing, etc **Encoder press**: change tune rate 1kHz->500Hz->100Hz->10Hz **Encoder dbl or long press**: choose stored message. Then Left to send repeatedly, Centre to send once, or Right to cancel	**Right single press**: change VFO mode: A, B, Split **Right Double press**: Select frequency preset. Then press Left (save VFO to preset), Right (load preset to VFO) or Centre (cancel) **Right long press**: swap VFO A and B frequencies

Configuration menu items:

1 Preset

Preset 1 to Preset 16: Enter frequency presets.

2 Messages

2.1 Message 1 to Message 12: Stored messages. First four are 100 characters, rest are 50 characters.
2.13 Interval: gap between message transmissions, in seconds
2.14 Repeats: number of message repeats, 1 to 99 or infinite

3 VFO

3.1 VFO mode: A, B or Split
3.2 VFO A: startup frequency
3.3 VFO B: startup frequency
3.4 Tune rate: startup rate, 1kHz, 500Hz, 100Hz or 10Hz
3.5 RIT: Receive incremental tuning
3.6 RIT rate: 1kHz to 1Hz
3.7 CW mode: CW or CW-R (reverse)
3.8 CW offset: default offset (700Hz)
3.9 Band: configure once at start-up

4 Keyer

4.1 Keyer mode: Straight, Iambic A/B, Ultimatic
4.2 Keyer speed: in words per minute
4.3 Keyer swap: swap paddle inputs
4.4 Keyer weight: alter dit:space ratio
4.5 Auto space: enable keyer autospacing (default OFF)
4.6 QSK: full or semi break in
4.7 Practice: enable practice mode, keys but no RF output
4.8 Sidetone frq: normally set to same as CW offset (700Hz)
4.9 Sidetone vol: volume, 0 to 99

5 Decoder

5.1 Noise blanker: period in ms
5.2 Speed Avg: for speed detection
5.3 Ampl. Avg: for ampl. detection
5.4 Enable Rx: enable Rx decoder
5.5 Enable Tx: enable Tx decoder
5.6 Enable Edit: enable CW decoding during menu editing (useful!)

6 Beacon

6.1 Beacon: On/Off enable beacon operation
6.2 Mode: CW or WSPR
6.3 Frequency: beacon frequency
6.4 Frame: frame duration in minutes
6.5 Start: minutes past the hour to start
6.6 WSPR call: Callsign to encode in WSPR message
6.7 WSPR locator 4-character Maidenhead square
6.8 WSPR power dBm power level to encode in WSPR message
6.9 Set time: set real time clock

7 Other

7.1 Dbl click: dbl click delay time in milliseconds (300)
7.2 Batt. show: enable battery icon on the display top right
7.3 Batt. Full: set millivolts for a "full" battery icon indication
7.4 Batt. set millivolts for each battery bar step level in the icon display
7.5 Cursor style: underline/flashing
7.6 Show S-meter: enable S-meter display on the screen
7.7 S-meter step: set value of each S-meter bar (in amplitude units)
7.8 Factory rst.: Factory reset, if set to value 17 – use with caution

8 Alignment

8.1 Alignment freq: frequency for BPF alignment
8.2 I-Q balance freq: audio frequency for I-Q balance adjustment (700Hz)
8.3 Phase adj Lo freq: audio freq for low audio phase adj (600Hz)
8.4 Phase adj Hi freq: audio freq for high audio phase adj (800Hz)
8.5 Ref. frq.: 27MHz synth reference frequency used for synthesiser
8.6 System frq.: 20MHz system clock frequency used for system timing
8.7 Peak BPF: do the actual BPF peaking alignment
8.8 I-Q bal.: do the I-Q balance adjustment
8.9 Phase Lo: do low audio phase adjustment
8.10 Phase Hi: do high audio phase adjustment
8.11 Cal ref. osc: do GPS calibration of 27MHz osc if GPS is connected
8.12 Cal sys. osc: do GPS calibration of 20MHz osc if GPS is connected
8.13 GPS data: show GPS satellite data if GPS is connected

9 Test equipment

9.1 Voltage: measure voltage
9.2 RF Power: measure RF power
9.3 Audio Ch.0: measure audio channel 0 amplitude
9.4 Audio Ch.1: measure audio channel 1 amplitude
9.5 Frequency: measure frequency
9.6 Signal gen.: enable and adjust signal generator output

Save settings!: save current VFO etc settings to be the default on power up; press left (select) to save, right to cancel

9. Resources

- For updates relating to this kit please visit the QRP Labs CW transceiver kit page http://qrp-labs.com/qcx
- For any questions regarding the assembly and operation of this kit please join the QRP Labs group, see http://qrp-labs.com/group for details

10. Document Revision History

1.00	02-Aug-2017	Initial revision for YOTA 2017
1.01	04-Aug-2017	Published draft version
1.02	18-Aug-2017	First official general release version
1.03	19-Aug-2017	Fixed some typos
1.04	21-Aug-2017	Corrected display description in section 4.2
1.05	31-Aug-2017	Added capacitor labels and resistor colour codes to parts list; added T1 diagram by Ed WA4MZS; minor corrections
1.06	25-Sep-2017	Added instruction to insert short end of the pin headers into PCB Added C47 and other minor corrections to parts list 6.4 IC pin count is anticlockwise (not clockwise) 6.4 table row 9, should be C43-47 (not 33-47) 3.17 The 1N4148 diode is at the board top left (not right) 5.14 GPS connector is at board right (not left) 4.2 Corrected description: top row is Rx VFO, bottom is Tx VFO 3.41 Corrected photo of 470uF which was reversed polarity
1.07	17-Oct-2017	Changed Q6 transistor to MPS751 (was MPS2907 in first kit batch) Corrected typo on p83 para 3 “in the event that” 3.3K R44 was mistakenly called R34 on parts list
1.08	23-Oct-2017	Swapped R19 and R25 in parts list and sections 3.22, 3.23 to match error on silkscreen (silkscreen prints for R19 and R25 were swapped). Note in section 3.41 that C47 can be mounted vertically above SN 500.

3.5 to 30 MHz Automatic Antenna Impedance Matching System

By

William Blodgett

Senior Project

ELECTRICAL ENGINEERING DEPARTMENT

California Polytechnic State University

San Luis Obispo

August 2012

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I could not have completed this project without their support.

- William Blodgett, KE6RKT, August 25, 2012

Abstract

Two-way communications using 3 to 30 MHz, high-frequency (HF) radio, also known as shortwave radio, provides worldwide coverage with no infrastructure required between stations. Amateur or "Ham" radio operators apply this advantage to long distance communications and to provide disaster relief communications. Optimum frequencies for long distance propagation are time-of-day variant necessitating an operating frequency range of 3 to 30 MHz. HF half wave dipole antenna length varies from 15 to 150 ft over this range. Fixed-length antenna impedance varies with frequency due to physical dimension dependent capacitance and inductance. Maximum transmitter to antenna power transfer occurs when the transmission line and antenna impedances are matched to the standard 50 Ω radio impedance. For transmitter to antenna impedance matching, antenna length can be adjusted depending on frequency. Alternately, an antenna tuner utilizing a variable inductance and capacitance matching network can compensate for the frequency dependent impedance of a fixed-length antenna. A dipole antenna designed for one frequency can function over a broader frequency range with a manually or automatically adjusted antenna tuner.

The project objective is to develop a microcontroller-based automatic tuner for a fixed-length dipole antenna. In high frequency transmission lines, an antenna impedance mismatch causes power reflection back to the transmitter. This corresponds to a greater than unity voltage standing wave ratio (VSWR), the ratio of maximum to minimum transmission line voltage amplitude. A directional wattmeter provides forward and reflected power values to an Atmel ATmega32 microcontroller, which calculates VSWR and adjusts a capacitor - inductor matching network using stepper motors to reduce VSWR to less than 1.5:1 (4% reflected power relative to incident). The antenna tuner system can match loads of up to 26:1 initial VSWR within a frequency range of 3.5 MHz to 30 MHz. The system was tested using a 100 W transceiver connected to a 90 ft, 450 Ω balanced transmission line and a 120 ft dipole antenna 30 ft above ground. Commercial antenna tuners have VSWR matching capabilities from 3:1 to 100:1. [1] [2] [3]

Chapter 1. Introduction

An antenna tuner, or impedance matching system (IMS), allows use of a fixed-length antenna on a range of frequencies wider than otherwise possible. Figure 1 shows a common setup for an antenna tuner. A fixed length wire dipole is usually mounted above ground using masts. The mounting height is at least one eight wavelength above ground for the lowest operating frequency. One eighth wavelength for 3.5 MHz is approximately 30 ft. If the antenna is closer to the ground, the feedpoint impedance approaches 0 Ω and the radiation pattern is directed upward instead of at the horizon, which is more desirable for long distance communication [4]. The dipole length is typically a half wavelength at the minimum frequency of the desired operating range. For example, if the minimum frequency is 3.5 MHz, the length is:

$$\left(\frac{1}{2}\right)\left(\frac{c}{f}\right) = \left(\frac{1}{2}\right)\frac{\left[3x10^8\,\frac{m}{s}\right]}{(3.5x10^6\ Hz)} = 43\ m\ or\ 140\ ft. \qquad (1)$$

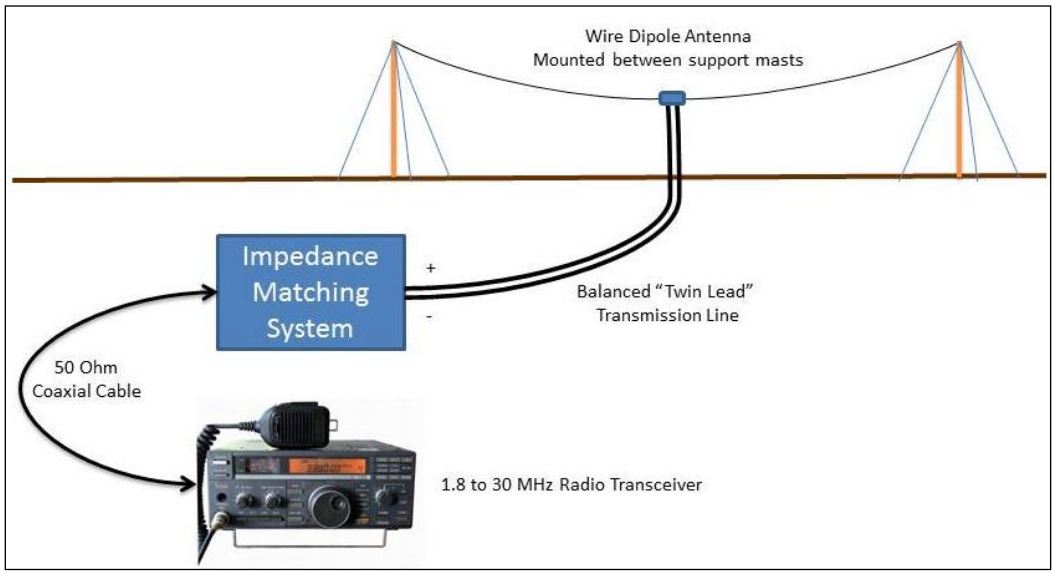

Figure 1 - Typical Impedance Matching System (IMS) Setup [5]

A typical IMS is shown in Figure 1. At power on, the IMS initializes in standby mode and measures and displays forward and reflected RF power. It calculates SWR if the RF input power is greater than 3 W. The power measurement range is 10 mW to 200 W. With the radio transceiver in receive mode, the RF power is less than 10 mW, therefore SWR calculation is not possible. Impedance matching is also important to maximize received signal strength. The user must transmit into the IMS and press the "Tune" button to initiate automatic impedance matching. The "Tune" button allows operator discretion for tuning initiation, although the IMS software could be modified to initiate tuning automatically if the VSWR exceeds 1.5:1 while

transmitting. During automatic impedance matching, the IMS calculates VSWR and actuates stepper motors to alternately adjust a variable capacitor and a variable inductor to reduce VSWR to less than 1.5:1. The IMS then returns to standby mode.

$$SWR = \frac{1 + \sqrt{\frac{P_R}{P_F}}}{1 - \sqrt{\frac{P_R}{P_F}}} \quad (2)$$

From equation (2), a VSWR of 1.5:1 corresponds to 4% reflected power. Reducing VSWR to less than 1.5:1 achieves an acceptable impedance match based on the manufacturer requirements of the transceiver used for IMS testing [6]. When transmit frequency is adjusted, an impedance mismatch will again exist requiring the operator to initiate another automatic tuning cycle to reduce VSWR.

Chapter 2. Background

Two-way communications using 3 MHz to 30 MHz, high-frequency (HF) radio, also known as shortwave radio, offers worldwide coverage with no required infrastructure between stations, making it useful for emergency communications. Worldwide range is possible due to HF radio wave refraction in the ionosphere known as skywave propagation. The ionosphere is the upper layer of Earth's atmosphere where solar radiation separates electrons from (ionizes) gas molecules. This ionized gas, known as plasma, is electrically conductive, causing incident radio waves to refract (bend) back to Earth. Earth's surface also reflects radio waves back to the ionosphere. Surface reflection and ionosphere refraction can occur multiple times along a radio signal path allowing the signal to propagate worldwide. The optimum frequency for long distance skywave propagation changes with time of day due to the sun's influence on the ionosphere. For transmitter to antenna impedance matching, antenna length can be adjusted depending on frequency. Half-wave dipole antenna length for HF radio varies from 15 ft to 150 ft.

A 140 ft dipole antenna normally resonates at 3.5 MHz. At the resonance frequency, inductive and capacitive reactances cancel. Away from resonance, the impedance changes and power reflections caused by impedance mismatch increase. An antenna tuner can render the antenna resonant to the transmitter and eliminate reflected power by providing compensation for the impedance mismatch. Matching is possible below the resonant frequency, but the voltage can exceed the dielectric strength of the transmission line or antenna tuner components. Table 1 shows possible voltages for a 100 ft "flat-top" antenna [5].

Modeled Data for a 100-ft Flat-Top Antenna

Freq (MHz)	*Antenna Impedance (Ω)*	*Input VSWR RG-213 Coax*	*Loss of 100 ft RG-213 Coax (dB)*	*Loss of 100 ft 450-Ω Line (dB)*	*Max Voltage RG-213 Coax at 1500 W*	*Max Voltage 450-Ω Line at 1500 W*
1.8	4.18 –*j* 1590	33.7	26.0	8.8	1507	10950
3.8	37.5 –*j* 354	16.7	5.7	0.5	1177	3231
7.1	447 +*j* 956	12.3	5.9	0.2	985	2001
10.1	2010 –*j* 2970	12.1	10.1	0.6	967	2911
14.1	87.6 –*j* 156	1.6	0.9	0.3	344	1632
18.1	1800 +*j* 1470	7.7	6.8	0.3	753	1600
21.1	461 –*j* 1250	4.6	3.2	0.1	585	828
24.9	155 +*j* 150	3.6	2.6	0.2	516	1328
28.4	2590 +*j* 772	6.7	9.4	0.5	703	1950

Notes

1) Antenna is a 100 ft long, 50 ft high, center-fed dipole over average ground, using coaxial (RG-213) or open-wire transmission lines. Each transmission line is 100 ft long.
2) Antenna impedance computed using *EZNEC-3* computer program using 499 segments and with the Real Ground model.
3) Note the extremely reactive impedance levels at many frequencies, but especially at 1.8 MHz. If this antenna is fed directly with RG-213 coax, the losses are unacceptably large on 160 meters, and undesirably high on most other bands also.
4) The RF voltage at 1.8 MHz for high-power operation with open-wire line is extremely high also, and would probably result in arcing either on the line itself, or more likely in the antenna tuner.

Table 1 - Loss Comparison of 450 Ω Balanced Line versus 50 Ω Coaxial Cable [5] Table 20.1. Used with permission. Copyright ARRL.

Voltage amplitude increases with reflection coefficient magnitude, |Γ|, between the antenna and transmission line and therefore increased voltage standing wave ratio (VSWR). Superposition of forward and reflected waves along the transmission line produces standing waves resulting in maximum and minimum voltage amplitude locations. VSWR is the ratio of the maximum to minimum transmission line voltage amplitude. ISWR is the corresponding ratio for current; its value is equivalent to VSWR. VSWR and ISWR are collectively referred to as SWR since they are equivalent. Equations (3) and (4) define Γ and SWR, respectively. [6]

$$\Gamma = \frac{Z_{Load} - Z_o}{Z_{Load} + Z_o} \qquad (3)$$

where Z_{Load} is the complex antenna (load) impedance and Z_o is the transmission line characteristic impedance. Z_o is typically assumed purely real for commercial transmission lines.

$$SWR = \frac{|V_{max}|}{|V_{max}|} = \frac{1+|\Gamma|}{1-|\Gamma|} \qquad (4)$$

Equation (5) defines reflection coefficient magnitude.

$$|\Gamma| = \sqrt{\frac{(R_{Load} - Z_o)^2 + (X_{Load})^2}{(R_{Load} + Z_o)^2 + (X_{Load})^2}} \qquad (5)$$

where X_{Load} is the antenna reactance and R_{Load} is the antenna resistance.

SWR can also be calculated from forward and reflected power using equation (2). The IMS employs forward and reflected power measurements to calculate SWR.

Equation (5) shows that if the antenna reactance is much greater than the antenna resistance and Z_o, the reflection coefficient, |Γ|, approaches 1. In equation (4), as |Γ| approaches 1, SWR approaches infinity, increasing maximum transmission line voltage.

Figure 2 shows that dipole antenna impedance becomes predominantly reactive below its resonant frequency of approximately 5 MHz. “In [Figure 2], the y-axis is calibrated in positive (inductive) series reactance up from the zero line, and negative (capacitive) series reactance in the downward direction. The numbers placed along the curve show the frequency in MHz. “ [4]

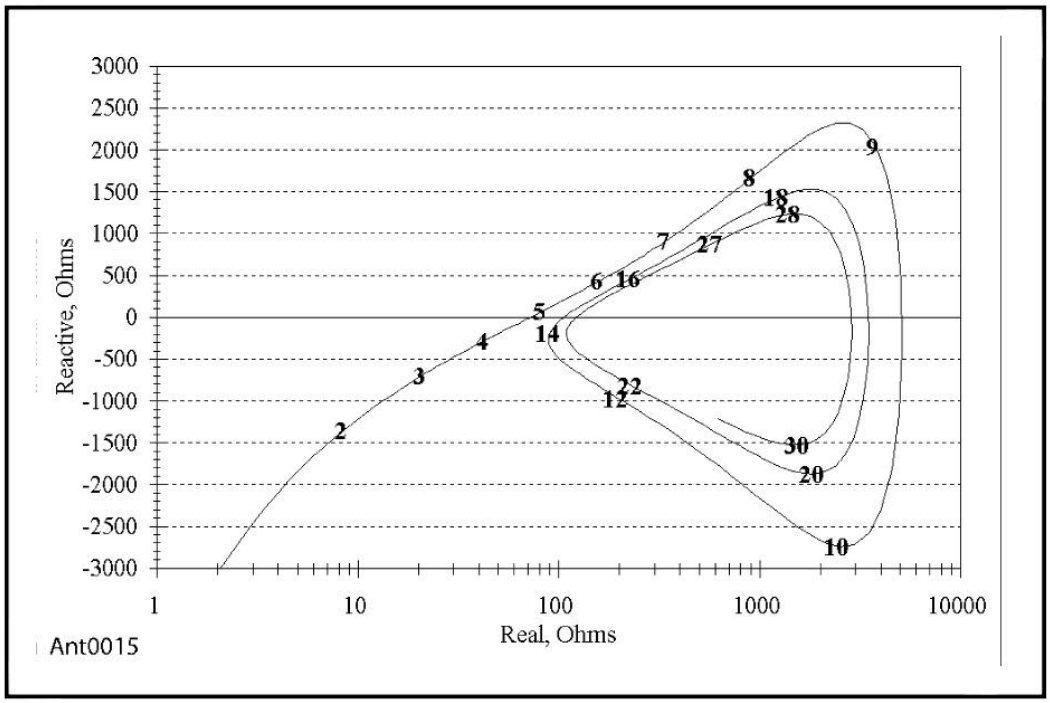

Figure 2 - Feedpoint impedance of a 100 ft long dipole antenna in free space. [4] Figure 2.10. Used with permission. Copyright ARRL.

In practice, radio operators typically use antenna tuners to match antennas for frequencies above the resonant frequency.

The IMS employs a balanced output design as opposed to unbalanced with unbalanced coaxial cable for both input and output. The problem with an unbalanced tuner is evident in Table 1 [7]. Terminating coaxial cable with highly mismatched loads results in undesirable power loss along the cable. A balanced transmission line, also called “twin lead,” conversely, exhibits lower loss when operated with mismatched loads. The IMS accepts an unbalanced line from the transceiver and internally converts it to a balanced line. The IMS to antenna transmission line is balanced. Since balanced transmission line is unshielded, unlike coaxial cable, it cannot be placed near other conductors. It must be spaced away from other conductors by at least three times the spacing between its conductors [7].

The IMS utilizes two L network matching circuit configurations. Compared to T or Pi networks, the L network uses only one combination of inductance and capacitance. This simplifies the microcontroller tuning algorithm. The disadvantage is that one L network cannot match all possible load impedances. Figure 3 shows two L network types with complimentary matching ranges on the Smith Chart. The IMS uses an RF switch to select one of the two L networks, allowing a wider matching range.

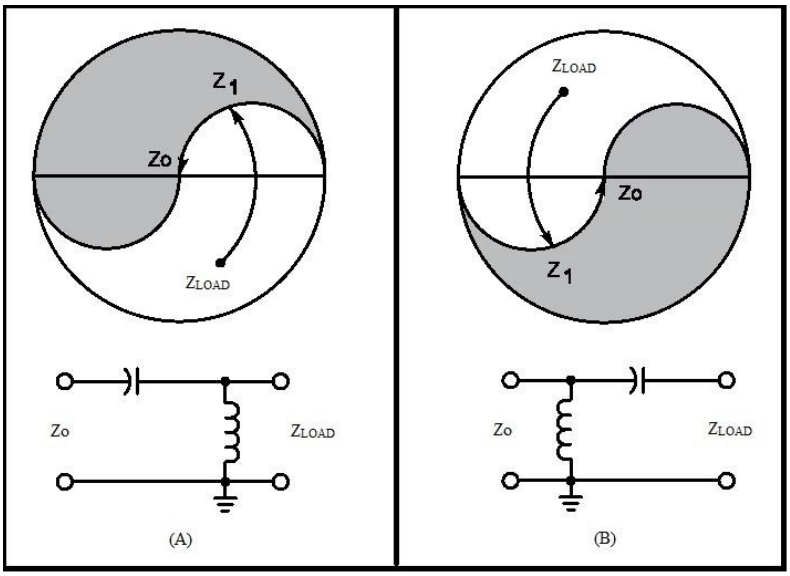

Figure 3 - Smith Chart operation and schematic for the two L network configurations [7] Figure 5.57. Used with permission; copyright ARRL. Modified from original.

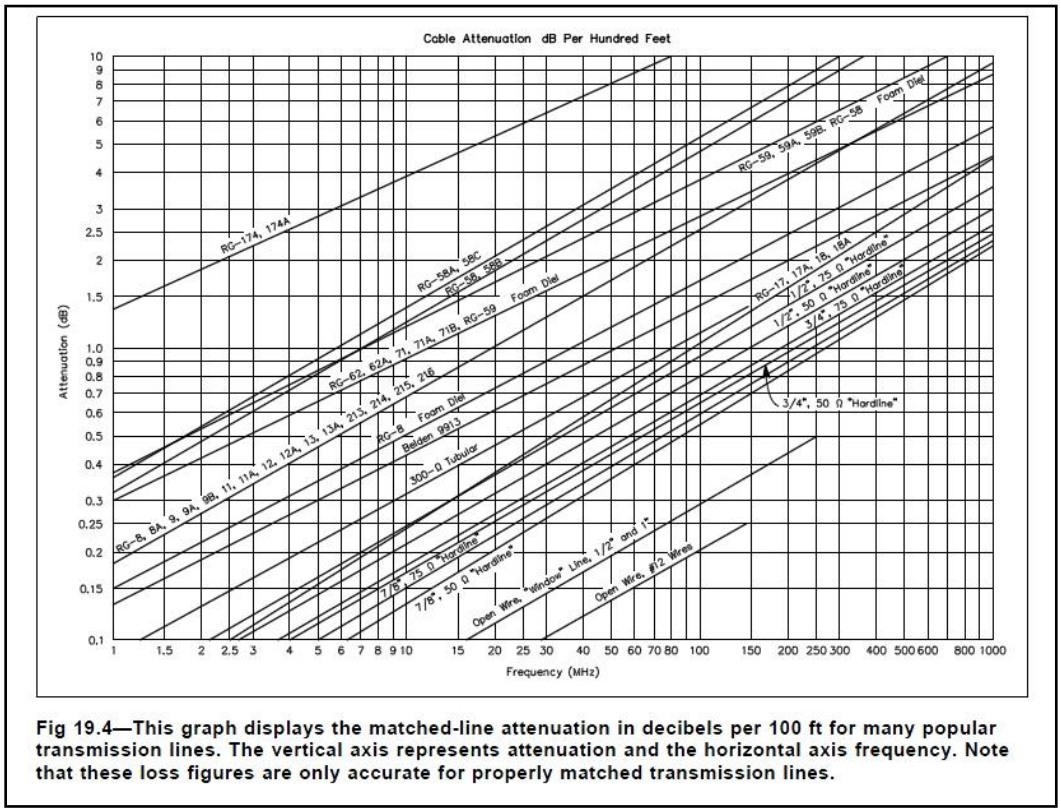

Fig 19.4—This graph displays the matched-line attenuation in decibels per 100 ft for many popular transmission lines. The vertical axis represents attenuation and the horizontal axis frequency. Note that these loss figures are only accurate for properly matched transmission lines.

Figure 4 – Transmission Line Loss Comparison [9]. Used with permission; copyright ARRL.

Figure 4 compares loss for several transmission line types. The two curves near the bottom represent two types of balanced transmission line, while upper curves represent coaxial cables. The loss for Open Wire "window" line, a twin parallel conductor balanced transmission line, is less than all types of coaxial cable for frequencies less than 200 MHz.

Chapter 3. Requirements

Table 2 defines specifications to achieve for the IMS design.

IMS REQUIREMENTS AND SPECIFICATIONS

Marketing Requirements	Engineering Specifications	Justification
1	1. Average setup time should not exceed 15 minutes	Required connections include antenna transmission line terminals, Impedance Matching System to radio transceiver coaxial cable, and 120 V AC power. (see Figure 5) The system should begin measuring RF power at power up. It will calculate SWR for RF input power greater than 3 W. To initiate automatic impedance matching, the user will press front panel button.
2	2. Parts cost should not exceed $1000	Similar impedance matching products range in market price from $200 to over $1000.Components required to build one unit must not exceed $1000. A preliminary parts search indicates total parts cost is less than $1000.
3	3. Can transfer at least 100 W of transmitter output power to a matched antenna continuously without damage.	All necessary parts are available and each can handle voltages and currents developed at 100 W RF power, the typical output power of most Amateur Radio HF transceivers.
1	4. Self-contained and portable with operating temperature range of 0° C to 70° C.	The unit will be housed in a portable (one person can carry) metal enclosure. The electronic components will be rated for commercial operating temperature range.
4	5. Antenna impedance matching range should allow matching to a dipole mounted at least 0.125 wavelength above ground and at least 0.5 wavelength long on the lowest operating frequency of 3.5 MHz.	A simple and effective portable HF antenna setup is a 140 ft wire dipole mounted 30 ft above ground. An antenna tuner should allow 3.5 MHz to 30 MHz operation on this antenna giving access to the most commonly used Amateur HF bands. One eighth wavelength for 3.5 MHz is approximately 30 ft If the antenna is closer to the ground, the feedpoint impedance approaches 0 Ω and the radiation pattern is directed upward instead of at the horizon, which is more desirable for long distance communication.

1, 3	6. The transmission line between the Impedance Matching System and the radio transceiver shall have a characteristic impedance of 50 Ω ±5%.	50 Ω is the standard antenna jack impedance on Amateur Radio HF transceivers and accompanying coaxial transmission lines. Coaxial transmission lines are unaffected when routed near metal; they can be used for outdoor unit to radio transceiver connections. The Impedance Matching System will automatically match the antenna impedance to 50 Ω.
4	7. The Impedance Matching System to antenna connection shall use terminals for a 300 to 600 Ω balanced transmission line.	Dipole, loop, and Yagi antennas are efficiently connected using a balanced transmission line with low loss in the 3.5 to 30 MHz range (see Figure 4) even when operated with unmatched loads. See Table 1, Columns 4 and 5
1,2	8. Operate using a standard 120 V AC wall outlet	Maximum power consumption occurs during matching attempts involving stepper motor operation. Stepper motor and control electronics power consumption should be below maximum of 1800 W for a 120 V AC outlet.
1	9. Must weigh less than 40 pounds	Similar products already on the market typically weigh less than 20 pounds. The weight must be less than 40 pounds for safe, single person lifting. The actual weight as built is 26 pounds.
1, 3	10. The system should warn the user if it is unable to reduce the standing wave ratio (SWR) to below 1.5:1.	Amateur Radio HF transceivers typically cannot operate with SWR greater than 2:1. Under these conditions, transceivers typically self-protect by reducing output power. The IC-706 HF transceiver manual [6] recommends less than 1.5:1 SWR. Transceivers have internal SWR meters to detect high SWR. Reflected power can increase thermal stress on the transceiver's internal power amplifier.
Marketing Requirements 1. Easy to set up and use 2. Affordable 3. Usable with most standard Amateur Radio HF transceivers 4. Works with many types of antennas		

Table 2 - IMS Requirements and Specifications Table [8], Chapter 3

Chapter 4. Design

The initial IMS was designed for an outdoor matching unit with an indoor status display and control unit. The outdoor unit simplifies routing the balanced transmission line away from metal objects as routing it inside a building is not required. Coaxial cable connects the IMS outdoor unit to the indoor radio position. This configuration exceeded cost and system complexity beyond the scope of the Senior Project. The outdoor unit requires an indoor status display unit with a separate microcontroller. Therefore, the design was revised to a self-contained indoor system with front panel status display and control. Indoor permanent installation is recommended. The location should allow balanced transmission line routing away from metal objects and ground.

Figure 5 defines the System Level Block Diagram. Inputs include electrical power and radio frequency power from the transceiver via coaxial cable. Outputs include status indicators and a balanced transmission line to the antenna. Table 3 provides additional details. While the coaxial cable and balanced transmission line are labeled input and output, they perform opposite functions when the transceiver is in receive mode.

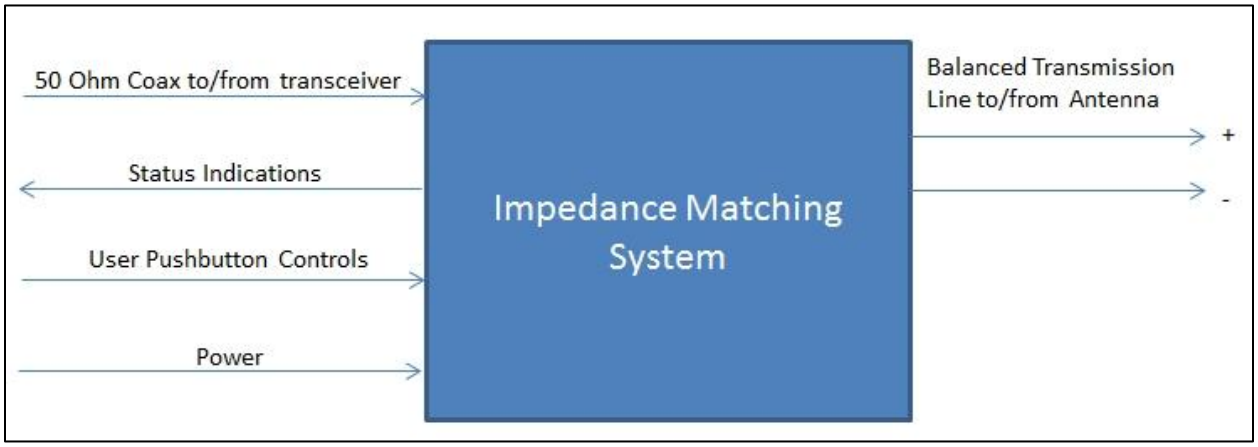

Figure 5 – System Level Block Diagram, Impedance Matching System

Table 3 – IMS System Level Functionality Table

Module:	**Impedance Matching System (IMS)**
Input	1. Power: 120 V AC at less than 10 A 2. 50 Ω Coaxial Cable from Transceiver: Carries 3.5 to 30 MHz transmit signal from the transceiver to the IMS. 3. User pushbuttons
Output	1. Balanced Transmission Line to Antenna: Carries transmit signal from the IMS to the antenna. 2. Status Indicators: An LCD provides user with current value of Voltage Standing Wave Ratio (VSWR), , forward and reflected power, inductor & capacitor relative positions (i.e. minimum and maximum inductance or capacitance), and RF Switch Position.
Functionality	Senses greater than 1.5:1 VSWR during transmit and adjusts internal reactance to match antenna transmission line impedance to the 50 Ω transceiver coaxial connection.

Figure 6 defines the IMS Subcomponent Block Diagram.

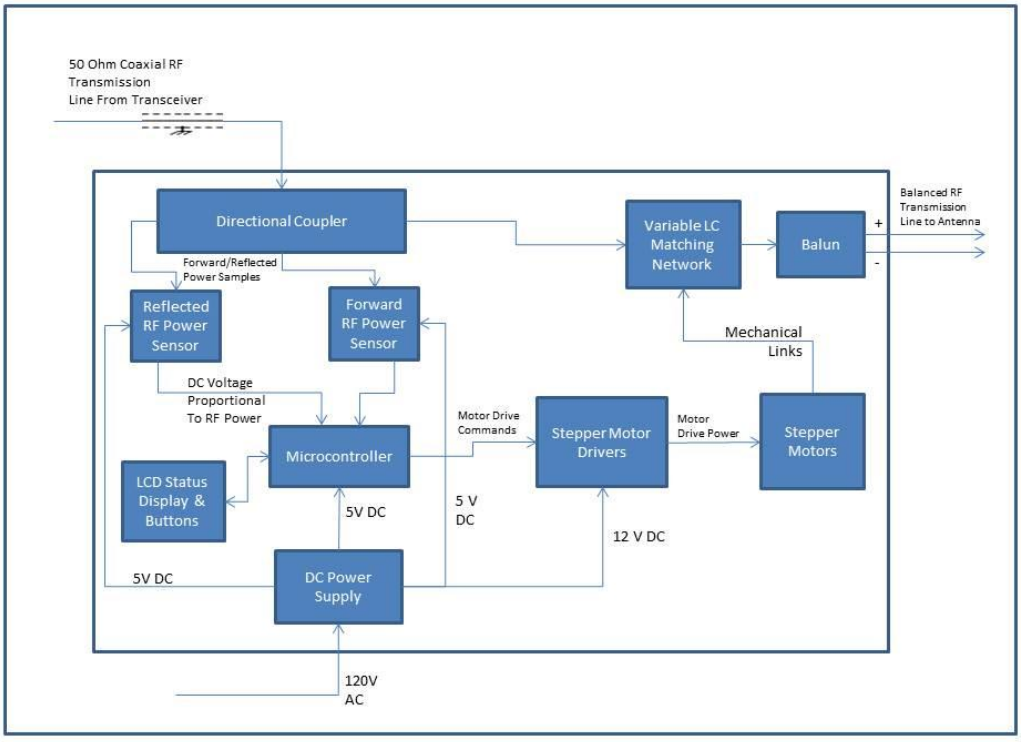

Figure 6 – IMS Subcomponent Block Diagram

Table 4 describes subcomponent inputs, outputs, and functionality.

Table 4 – IMS Subcomponent Level Functionality Table

Module		Description
Forward/Reflected RF Power Sensor	Input	1. Radio Frequency Power from Directional Coupler. 2. 5 V DC Power
	Output	1. DC Voltage Proportional to RF Power
	Functionality	Produces two DC voltage outputs: one proportional to the forward RF power and one proportional to reflected RF power.
Directional Coupler	Input	1. RF Power from Transceiver
	Output	1. RF Power to LC matching network 2. Forward Power Sample to Forward RF Power Sensor 3. Reflected Power Sample to Reflected RF Power Sensor
	Functionality	Provides 30 dB reduced samples of the forward and reflected RF power to allow RF power sensor power measurement
Balun	Input	4. RF power via unbalanced coaxial transmission line
	Output	1. RF power via balanced transmission line
	Functionality	A Balun is a transformer that converts an unbalanced line to a balanced line.
Variable LC Matching Network	Input	1. RF power from Directional Coupler 2. Mechanical Actuation from Stepper Motors
	Output	1. RF Power to Balun
	Functionality	The LC Matching network is a shunt variable inductor and a series variable capacitor that each provide the necessary reactance to achieve an impedance match. A single pole, double throw RF switch places the shunt inductor on either side of the capacitor to match a wider range of impedances.
Microcontroller	Input	1. DC Voltage Proportional to RF Power 2. DC Power 3. User Pushbutton Controls
	Output	1. Stepper Motor Drive Commands 2. Data to LCD Status Display
	Functionality	The microcontroller measures DC voltage from the RF power meter and issues motor drive commands to adjust the LC network for an impedance match. It also provides serial data to display RF power level and tuning status.
DC Power Supply	Input	1. 120 V AC power
	Output	1. DC power: 12 V at 2 A and 5 V at 1 A.
	Functionality	Converts 120 V AC to 12 V and 5 V DC power for the circuitry and stepper motors.
Stepper Motor Drivers (3)	Input	1. Motor Drive Commands from Microcontroller 2. DC Power
	Output	1. Motor Drive voltage
	Functionality	Provides motor power based on commands received from the microcontroller.
L and C Stepper Motors	Input	1. Motor Drive Power
	Output	1. Mechanical Rotation
	Functionality	There are three stepper motors: The L and C motors adjust the variable inductor and capacitor, and a linear actuator motor controls the RF switch in the matching network
LCD Status Display	Input	1. Data from microcontroller 2. DC Power
	Output	1. Display information
	Functionality	The Status Display Unit displays forward and reflected power and tuning status received from the microcontroller.

Directional Coupler

A directional coupler is an RF power measurement device with four ports: RF input, RF output, forward coupled (or sampled), and reflected (or reverse) coupled. A directional coupler samples a portion (coupling factor) of RF power in the forward and reverse directions. The coupling factor (separate values for forward and reverse directions) is the ratio of actual to sampled RF power. The RF input and output ports are connected by a main (through) line and the forward and reflected coupled ports are connected by a coupled line. A portion of the main line RF power is diverted to the coupled line where it is attenuated by a fixed value known as the coupling factor (ratio of actual to sampled RF power). The coupling factor is high, ~1000 or 30 dB, to minimize main line power loss. The forward coupled port power is coupled in phase with the main line forward power and out of phase with main line reflected power, further attenuating reflected power by a factor known as directivity. The opposite case is true for the reflected coupled port. The directivity is the power ratio of the forward to reflected coupled ports with the through port terminated in a matched ($\Gamma = 0$) load. Directivity is ideally infinite since there should be no reflected coupled power for a matched load at the through port. The directional coupler allows transmission line forward and reflected power measurement using sensitive low power electronics while minimizing main line power loss.

The IMS employs a directional coupler for forward and reflected power measurement and SWR calculation. The directional coupler enables 200 W power measurements using the AD8307 logarithmic detector IC, which has a +17 dBm (50 mW) maximum power input. [10] A directional coupler design by W. T. Kaune [11] shown in Figure 7 utilizes inexpensive components.

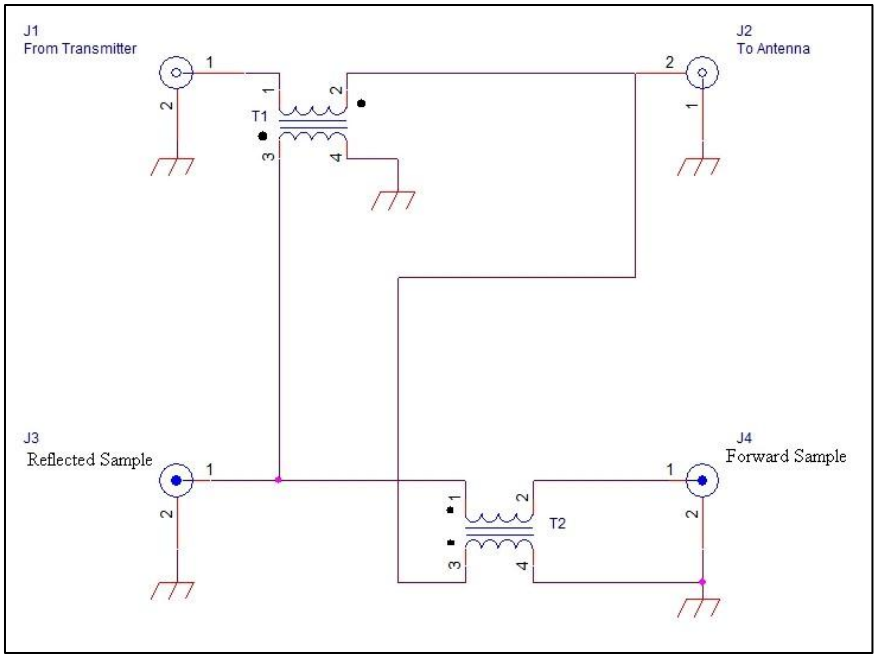

Figure 7 - Directional Coupler Circuit Diagram [11]

Ferrite toroids wound with AWG 26 wire and surrounding two 2" sections of RG-8 50 Ω coaxial cable form the coupling transformers. The shield conductor of the coaxial cable through the toroids prevents capacitive coupling between the transmission line and windings. Double sided circuit board provides additional shielding to prevent unwanted RF coupling.

Directivity is the ratio of forward to reflected coupled power when the directional coupler is terminated with matched (Γ=0) loads. Ideally, directivity approaches infinity since the reflected coupled port power should be zero for a matched load. Commercial HF directional couplers offer directivities from 15 to 44 dB. [12] [13] Kaune [11] claims a directivity of 35 dB at 3.5 MHz and 28 dB at 30 MHz.

RF power levels, in W and dBm (dB referenced to 1mW) are defined in equations (6) and (7).

$$P(dBm) = 10\,log_{10}[P(mW)] = 10\log_{10}[1000([P\,(W)])] \quad (6)$$

$$P\,(W) = \frac{10^{\left[\frac{P(dBm)}{10}\right]}}{1000\,\frac{mW}{W}} \quad (7)$$

The directional coupler provides a forward coupled power output 30 dB less than the forward transmitter power. For 100 W (50 dBm from equation (6)) forward power, the forward coupled power should be 20 dBm. With 28 dB of directivity, the reflected coupled port is:

20 dBm -28 dB = -8 dBm = 0.16 mW (8)

Using equation (2) to calculate SWR from forward and reflected power, the SWR measured using this directional coupler is 1.08 and 1.03 for 28 dB and 35 dB of directivity, respectively.

Power Supply

Four devices require power: the microcontroller, the power sensors, the LCD, and the stepper motor drivers. The stepper motor drivers require 12 ± 1 V at 1 A. Table 5 defines the 5 V supply requirements.

5 V DC Power Supply Requirements		
Device	Voltage Range, V	Maximum Current, mA
ATmega32 Microcontroller [14]	4.5 to 5.5	200
AD8307 Power Sensor, Quantity 2 [10]	2.7 to 5.5	20
LCD [15]	4.7 to 5.5	165
Power Supply Requirement	4.7 to 5.5	385

Table 5 - 5 V DC Power Supply Requirements

The supply current capability is increased to 1 A for the 5 V supply and 2 A for the 12 V supply for improved reliability. A power supply providing half its current capability will outlast one supplying full capacity.

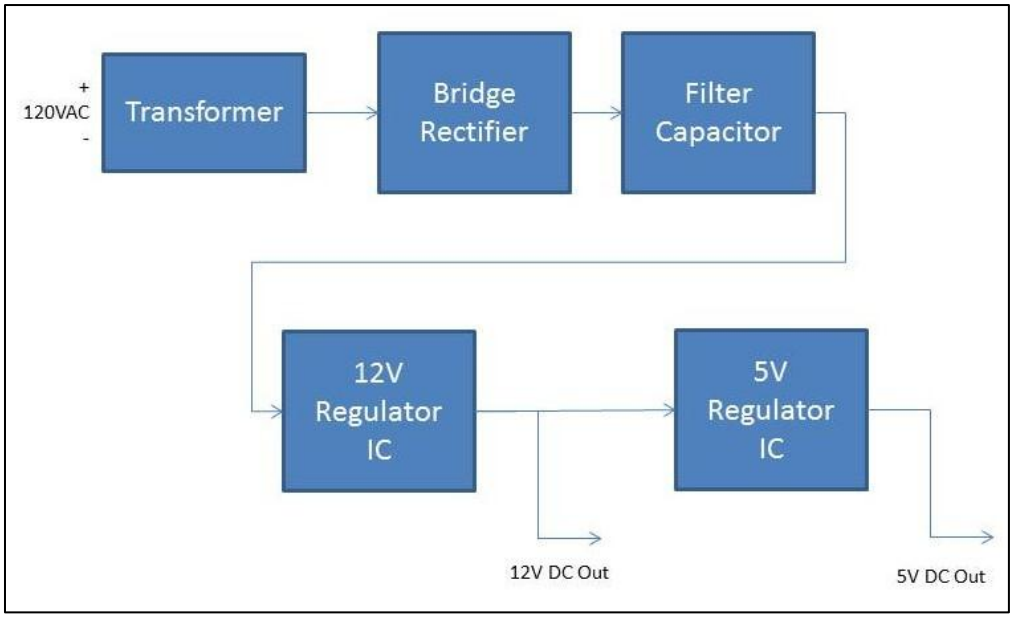

Figure 8 - Power Supply Block Diagram

Figure 8 shows the power supply conceptual block diagram. A 120 V/25.2 V 2 A center tapped transformer was selected. The center tap provides (25.2 V)/2 or 12.6 V AC RMS. Though only a 2 A rectifier was required, a 4 A bridge rectifier was selected for reliability. Fixed voltage linear

three terminal regulators were selected: ST Microelectronics L78S12 12 V, 2 A regulator [16] and Texas Instruments TL780-05 5 V, 1.5 A regulator [17]. Both feature short circuit and thermal overload protection.

The filter capacitance was determined as follows.

The transformer peak voltage output is:

$$V_{peak=}\sqrt{2}V_{RMS} = \sqrt{2}\,(12.6\,V) = 17.8\,V \quad (9)$$

Assuming 2 x 0.7 V diode voltage drop, the bridge rectifier output is 17.8 V - 1.4 V = 16.4 V

The L78S12 voltage regulator IC has a required minimum input voltage of 14.5 V. Thus, the maximum voltage ripple is 16.4 V – 14.5 V = 1.9 V.

Assuming maximum supply current of 2 A; equation (10) gives the required capacitance.

$$C = \frac{I}{2fV_{ripple}} = \frac{2\,A}{2(60\,Hz)1.9\,V} = 8800\,\mu F \quad (10)$$

Thus, two 4700 µF capacitors in parallel form the filter capacitor.

Balun

A balun transforms a balanced line to an unbalanced line. The IMS input and matching network are unbalanced, meaning one terminal is at ground potential. The balun at the IMS output enables balanced transmission line use for the IMS to antenna connection. The ARRL Handbook [7] provides a suitable balun design shown in Figure 9.

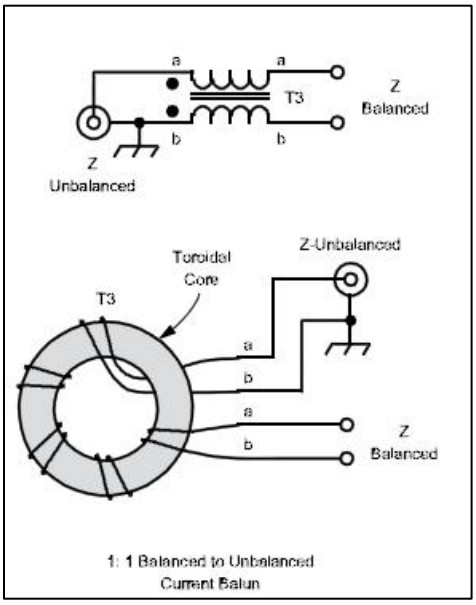

Figure 9 - 1:1 Balun Design [5] Figure 20.23A. Used with permission; copyright ARRL

The design calls for 12 bifilar turns of AWG 14 enamel insulated wire on a 2.4 inch outer diameter mix type 31 or 43 ferrite toroid. A 1:1 balun does not transform impedance. With the balanced terminals connected to a 50 Ω load, the unbalanced terminal impedance is also 50 Ω.

RF Power Sensor

An RF power sensor design by W. T. Kaune [11] , shown in Figure 10, utilizes the Analog Devices AD8307 Log Amp Detector IC, which outputs a DC voltage proportional to RF input power from -75 dBm to +17 dBm. Equation (11) yields this relation. [10]

$$V_{out} = 0.025\big(P_{in}(dBm)\big) + 2 \quad (11)$$

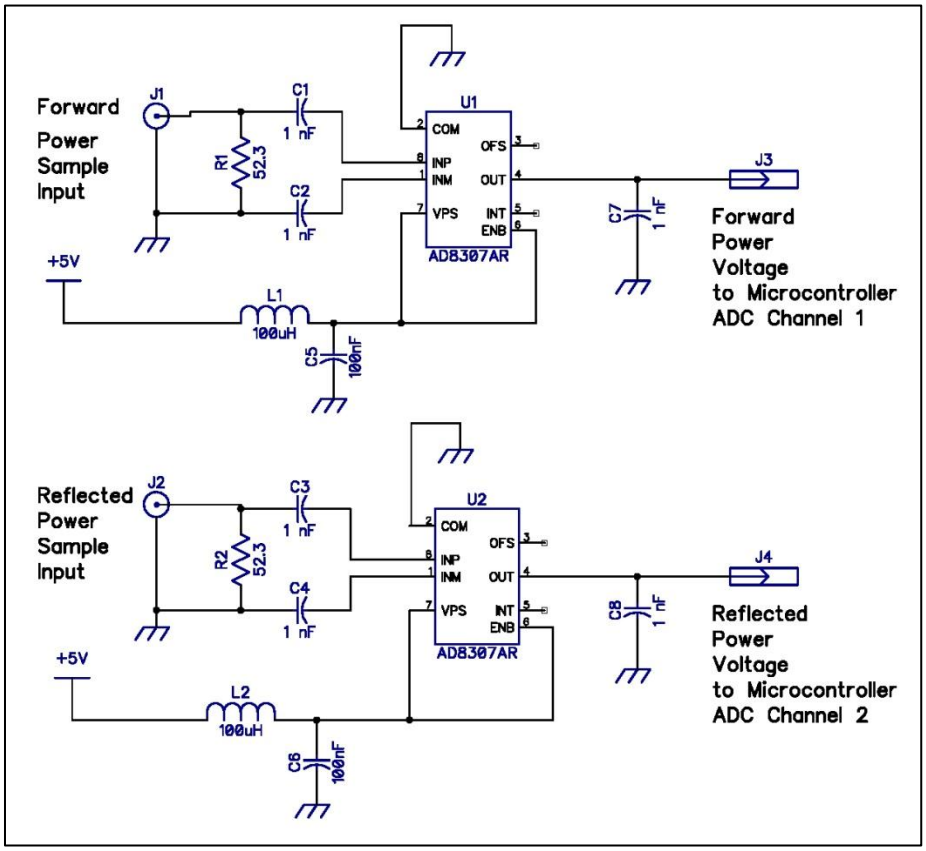

Figure 10 - RF Power Sensor Design [11]

Microcontroller

The Atmel Atmega32 microcontroller IC was selected due to previous experience in CPE-329. It features 32 programmable I/O pins, 32 kB non-volatile program memory, and an eight channel, 10 bit analog to digital converter (ADC). The internal ADC allows direct connection of the RF sensor output.

LCD

A two line by 16 character LCD displays forward & reflected power in W, SWR, inductor & capacitor positions, and RF switch position. The LCD requires 11 connections to the microcontroller.

Stepper Motors

The stepper motors were the largest available from Sparkfun.com, with 125 oz.-inches of torque [18]. They are rated at 2 A at 3 V. The torque required to rotate the capacitor and inductor was unknown since a measurement device was not available. Therefore, motor adequacy was determined by actuating the capacitor and inductor components. A third stepper motor is a linear actuator that controls the matching network RF switch [19].

Novel/Innovative Features

The matching network features a switch to select one of two possible L network configurations: series C shunt L and shunt L series C. The switch places the shunt inductor on either the source or load side of the capacitor. One L network configuration can only match half of all possible loads. The switch enables matching all possible loads by allowing both L network configurations. While the switch could have been implemented using a relay, this option requires power to maintain the relay in its default position. Latching relays, which remain in either set position with power removed are available, but a linear actuator stepper motor was already on hand. The linear actuator positions a rod in one direction or in the opposite direction. Its rotor shaft is a threaded tube with a threaded rod inserted. If the threaded rod is prevented from spinning with the rotor shaft, the rotor threads the rod in one direction depending on the rotation direction. The rod is attached to the switch's moving contact positioning it to connect to either stationary contact.

Stepper Motor Driver

The "Big Easy Driver" [20] operates bipolar stepper motors up to 2 A and 8 – 30 V. Microcontroller interfacing requires three output pins to control three functions: Step, Dir, and Enable. "Enable" turns on the driver and applies motor drive current. "Step" rotates the motor one step for each rising edge. "Dir" defines "Step" rotation direction; clockwise or counterclockwise.

Chapter 5. Test Plans

Directional Coupler

Test 1:
The directional coupler should not present greater than 1.5:1 SWR at its RF input with its outputs terminated in 50 Ω (Γ = 0) loads. Measure RF input SWR using a vector network analyzer over the 3.5 to 30 MHz frequency range with all outputs terminated in 50 Ω.

Test 2:
The coupling factor (ratio of actual to sampled RF power) should be 30±1 dB [11]. Connect an RF signal generator set to 0 dBm to the RF input port and measure the forward coupled port output power using a spectrum analyzer to determine the coupling factor. Ideally, the coupling factor should not vary by more than 1 dB from 3.5 to 30 MHz.

Test 3:
Minimum directivity, equation (12) below, is 28 dB [11]. Connect an RF signal generator set to 0 dBm to the RF input port and terminate the RF output port in 50 Ω. Measure the RF power at the forward and reflected coupled ports using a spectrum analyzer.

$$Directivity, (dB) = [fwd.coupled\ pwr, (dBm)] - [refl.coupled\ pwr, (dBm)] \qquad (12)$$

Minimum directivity is 28 dB over 3.5 to 30 MHz range.

RF Power Sensor

Verify that the DC output voltage conforms to equation (13) from -75 dBm to +17 dBm.

$$V_{out}, (V) = A\big(P_{in}, (dBm)\big) + B, (V) \qquad (13)$$

Where A = 23 – 27 mV/dB and B = 1.9 – 2.2 V [10]

Connect an RF signal generator to the sensor RF input and measure the DC output voltage for power levels from -75 dBm to +17 dBm at 3.5, 15, and 30 MHz.

DC Power Supply

Test power supply for output voltage regulation under no load and full load conditions. Full load is 0.5 A and 1 A for the 5 V and 12 V supplies, respectively. Apply 120 V AC input power and measure the 5 V and 12 V supply output voltages for the no load condition. Connect a 10 Ω, >2.5 W resistor to the 5 V output and a 12 Ω, >12 W resistor to the 12 V output. Measure the 5 V and 12 V supply output voltages for the full load condition. The no load and full load voltages should not exceed the voltage regulator datasheet specifications of 0.05 V and 0.6 V for the 5 V and 12 V supplies, respectively. [16] [17].

Stepper Motors

Rotate the variable capacitor and inductor using the stepper motors. Rotation will not occur if the required torque exceeds the 125 oz/inch motor rating.

LC Matching Network

Connect a 50 Ω dummy load to the matching network output and a 3.5-30 MHz transmitter and SWR meter to the input. Manually tune the matching network from 3.5 to 30 MHz. The matching network should achieve maximum 1.5 SWR matching from 3.5 to 30 MHz.

Microcontroller

The microcontroller IC is tested using the Atmel STK-600 microcontroller development board. The USB powered STK-600 has a plug in socket for the microcontroller IC. The microcontroller pins are routed to pins on the board which can be connected to onboard LED's and switches. The STK-600 can also program the microcontroller via the USB connection. The microcontroller is programmed to illuminate the LED's to verify all the port pins. The IMS microcontroller circuit board is tested with the same LED blink program.

LCD

A program displaying upward counting numbers verifies LCD functionality.

Automatic Impedance Matching

The impedance of a 120 ft dipole antenna connected using 90 ft of 450 Ω balanced transmission line is measured from 3.5 to 30 MHz using a vector network analyzer. With the IMS connected to the dipole, attempt automatic impedance matching for center and edge frequencies in all Amateur radio bands from 3.5 – 30 MHz. Table 7 lists the bands.

Table 6 – Amateur Radio HF Bands

Amateur Radio HF Bands	
Approximate Wavelength, M	Frequency, MHz
80	3.5 – 4.0
40	7.0 – 7.3
30	10.1 – 10.15
20	14.0 – 14.35
17	18.068 – 18.168
15	21.0 – 21.450
12	24.89 – 24.99
10	28.0 – 29.7

Dipole impedance values for frequencies where maximum 1.5 SWR matching is achieved are plotted on a Smith chart. The impedance range for maximum 1.5 SWR matching is determined using the Smith chart.

Chapter 6. Development and Construction

Chassis

The main chassis was salvaged from a General Electric Mobile Radio MastrII® Base Station power supply [21]. The chassis provides interior dimensions of 15.5" x 10.25" x 6". All interior components were removed. For easier drilling, the steel front and rear panels were replaced with 0.125" thick aluminum sheets matching the dimensions of the original panels. A 14.5" x 9.25" aluminum panel was bolted to the interior bottom to serve as an attachment point for all IMS components. Mounting holes were drilled and tapped in this panel allowing all component mounting screws and bolts to be threaded directly into the sheet. Rubber bumpers were bolted to the chassis underside to prevent sliding and scratching of tables and counter tops.

Directional Coupler

Figure 11 shows the completed directional coupler. Except for copper tape used to secure the circuit board shields, it was built as specified by Kaune [11].

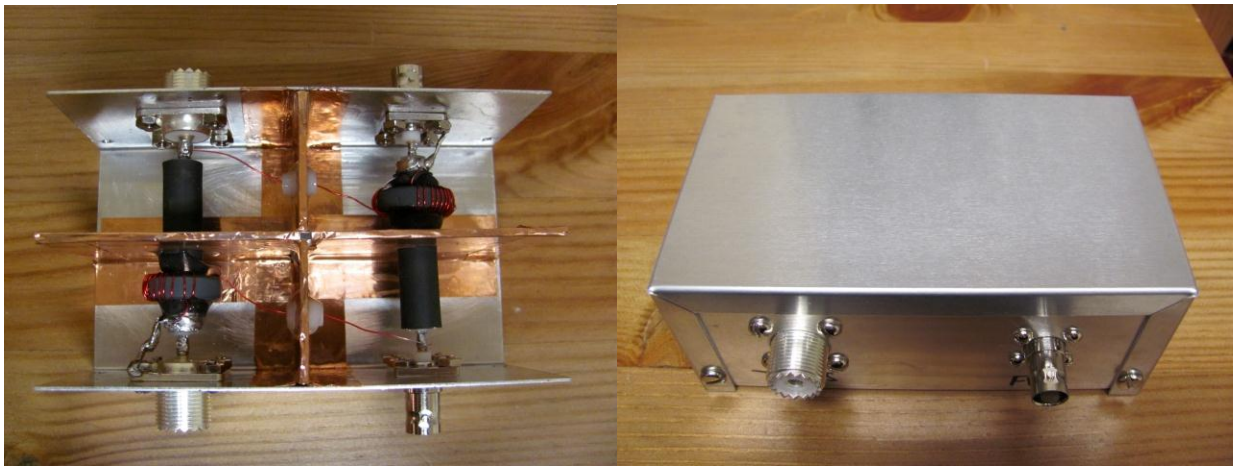

Figure 11 – Directional Coupler

Power Supply

All power supply components were grouped in the rear left corner to minimize space. The power switch and fuse holder were salvaged from the General Electric power supply [21]. The power supply circuit board, shown in Figure 12, was mounted to the rear panel with the voltage regulators bolted to the panel for improved heat dissipation.

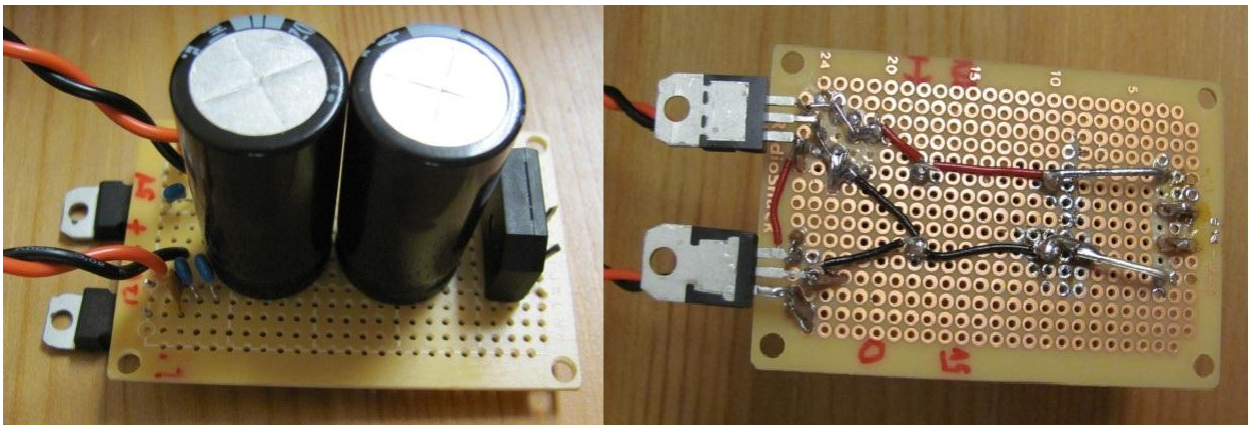

Figure 12 - Power Supply Circuit Board

Figure 13 shows the complete power supply. The power switch, fuse holder, and AC power cord plug (male IEC plug) were mounted to the IMS rear panel. Heat shrink tubing and liquid electrical tape on 120 V AC wires and terminals mitigate electrical shock and short circuit hazards. The transformer is bolted to the chassis bottom plate.

Figure 13 - IMS Power Supply

Matching Network

Figure 14 shows the complete matching network. The variable capacitor was salvaged from a manual antenna tuner and the inductor was purchased new. The inductor and capacitor motor mounts were fabricated from an aluminum project box. Motor positioning was critical to ensure alignment with the inductor and capacitor rotation axis preventing wobble during rotation. One inductor terminal is grounded to the chassis at its mounting bracket. The other terminal connects to the movable contact of the RF switch.

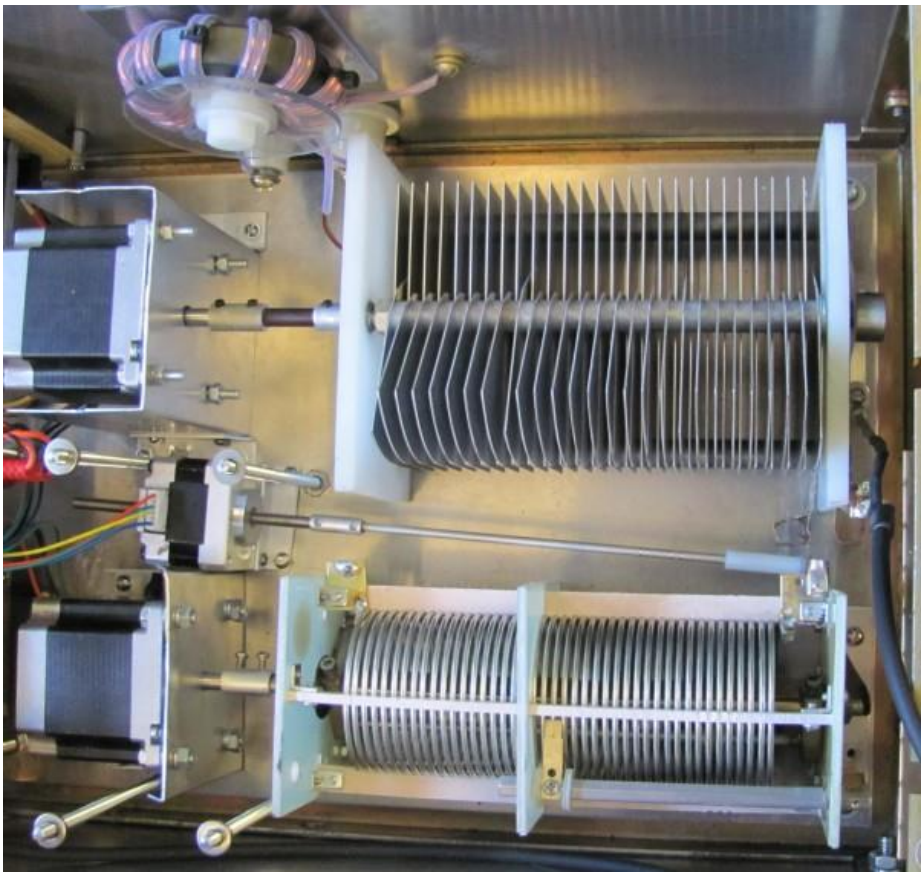

Figure 14 - IMS Matching Network

Figure 15 shows the RF switch built for the matching network.

Figure 15 - Matching Network RF Switch

The switch has two stationary contacts connecting to either the load side or source side of the capacitor. A threaded nylon spacer connects the push rod to the movable contact insulating the contact from the grounded rod.

Balun

The 1:1 balun was built as shown in Figure 9. Figure 16 shows the completed balun.

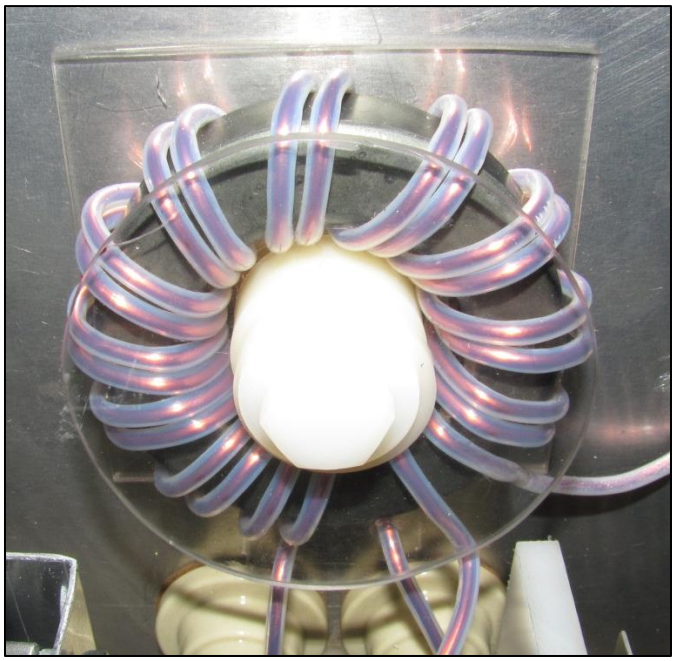

Figure 16 - IMS Balun

The wire is Teflon insulated to prevent arcing. A square polycarbonate sheet insulates the balun from the rear panel and a round sheet serves as a washer to hold the balun in place. Nylon washers, a bolt, and a nut fasten the balun to the rear panel.

RF Power Sensors

The two RF power sensors, forward and reflected, were built on pre-fabricated PCB's from Far Circuits [22]. A shielded box was constructed using copper clad PCB to minimize RF interference to the sensors. Figure 17 shows the completed RF power sensors in the shielded box. The box lid attaches with four screws and is removed for the photo.

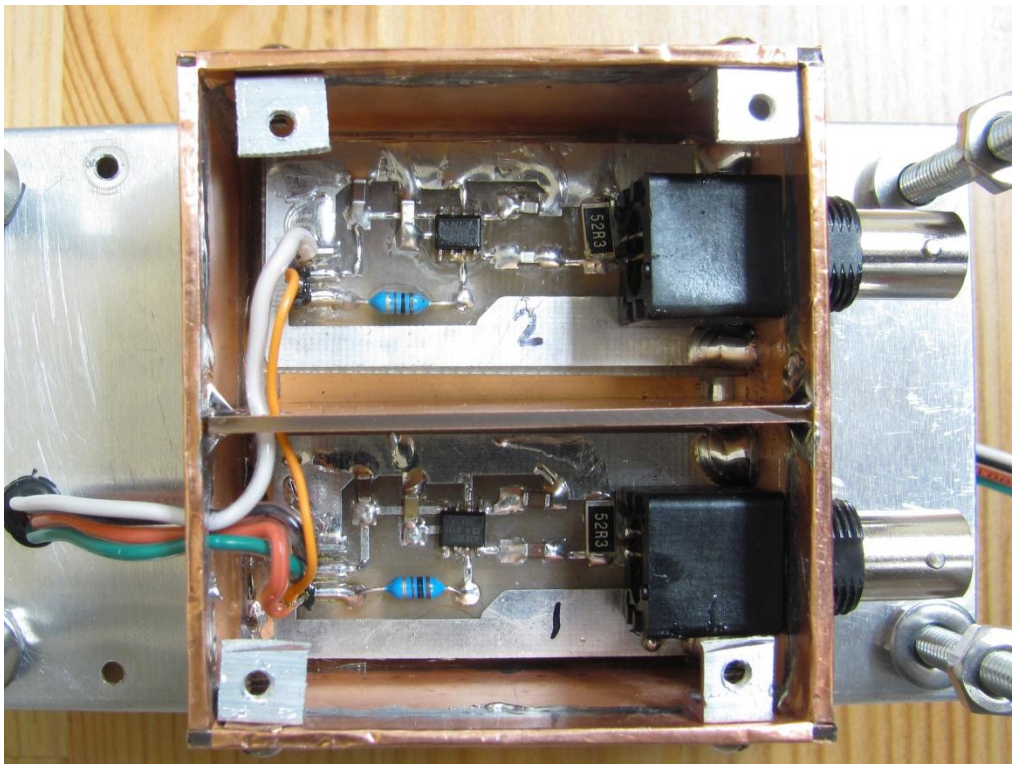

Figure 17 - RF Power Sensors

The RF power sensor module is attached to the bottom of the microcontroller module using screws and nuts. The power, ground, and sensor voltage wires route inside the microcontroller module where they are soldered in place. The microcontroller and RF sensor modules are removable from the IMS as one unit.

Microcontroller Module

The microcontroller circuit was built on a prototyping perf board. A 40 pin IC socket was soldered to the board to facilitate microcontroller replacement. Figure 18 shows the bottom of the microcontroller circuit board and the solder connections.

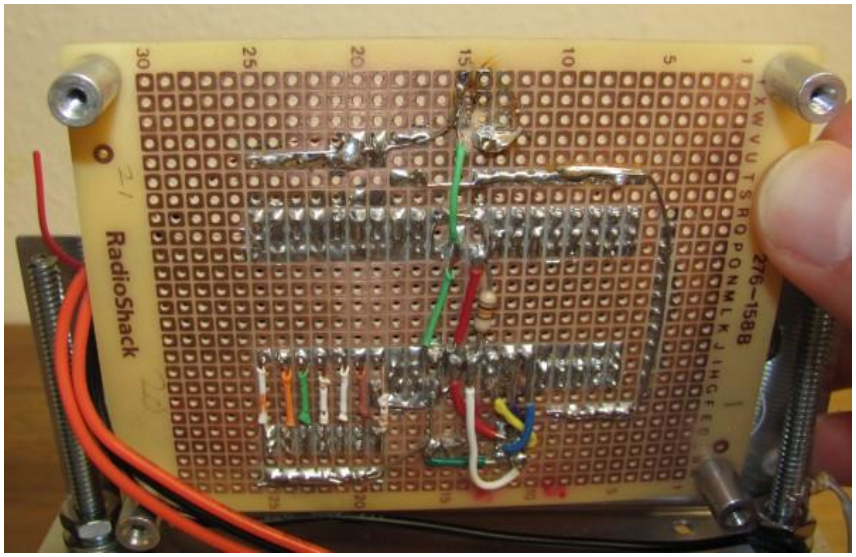

Figure 18 - Bottom of Microcontroller Board

Header pins were wired to the microcontroller I/O pins providing connection points for system development. Permanent I/O connections use wire wraps to the header pins.

To allow software modification, a 2 row by 3 pin male header plug was installed on the circuit board and connected to the microcontroller "In System Programming" (ISP) interface. A computer USB interface allows microcontroller software programing. Figure 19 shows the completed microcontroller module.

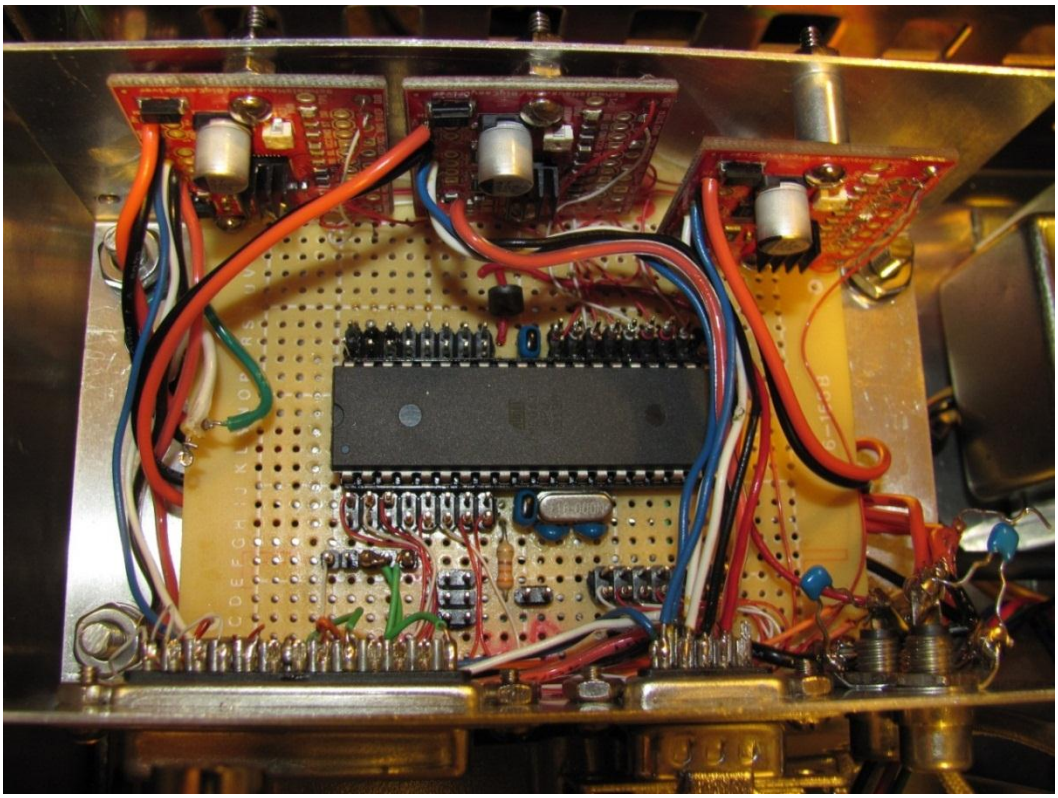

Figure 19 - Microcontroller Module

The microcontroller board is attached to the floor of an aluminum project box using screws and 0.5” standoffs. The three stepper motor drivers are mounted along one wall in the box. Connectors installed on the box interface the module to the IMS. A DB25 connector connects to the IMS front panel LCD and buttons. A DB15 connector connects all three stepper motors. 12 V and 5 V DC power are supplied to the module via RCA connectors. Table 7 lists all connections made to the microcontroller pins.

Atmega32 Microcontroller Pin Connections				
Pin Number	Pin Name	Input/Output	Destination	Use
1	Port B - 0	Output	Front Panel LCD Via DB25 Pin 24	LCD Register Select (RS)
2	Port B - 1	Output	Front Panel LCD Via DB25 Pin 23	LCD Read/Write Select (R/W)
3	Port B - 2	Output	Front Panel LCD Via DB25 Pin 22	LCD Enable ('E)
4	Port B - 3	Input	Front Panel Via DB25 Pin 12	"Select" Button
5	Port B - 4	Input	Front Panel Via DB25 Pin 11 and ISP SS Pin	"Select + " Button and ISP
6	Port B - 5	Input	Front Panel Via DB25 Pin 10 and ISP MOSI Pin	"Select - " Button and ISP
7	Port B - 6	Input	Front Panel Via DB25 Pin 9 and and ISP MISO Pin	"Tune Start/Stop" Button and ISP
8	Port B - 7	Output	Inductor Motor Driver DIR Pin and ISP SCK Pin	Inductor Motor Direction Select and ISP
9	RESET	Input	MCU Circuit Board Reset Pin	Resets Microcontroller
10	VCC	-	`+5 V DC Supply	Microcontroller DC Power Input
11	GND	-	Ground	
12	XTAL2	-	16 MHz Crystal	Microcontroller Clock Reference
13	XTAL1	-	16 MHz Crystal	Microcontroller Clock Reference
14	Port D - 0	Output	Front Panel LCD via DB25 Pin 21	LCD Data Buss 0 (DB0)
15	Port D - 1	Output	Front Panel LCD via DB25 Pin 20	LCD Data Buss 1 (DB1)
16	Port D - 2	Output	Front Panel LCD via DB25 Pin 19	LCD Data Buss 2 (DB2)
17	Port D - 3	Output	Front Panel LCD via DB25 Pin 18	LCD Data Buss 3 (DB3)
18	Port D - 4	Output	Front Panel LCD via DB25 Pin 17	LCD Data Buss 4 (DB4)
19	Port D - 5	Output	Front Panel LCD via DB25 Pin 16	LCD Data Buss 5 (DB5)
20	Port D - 6	Output	Front Panel LCD via DB25 Pin 15	LCD Data Buss 6 (DB6)
21	Port D - 7	Output	Front Panel LCD via DB25 Pin 14	LCD Data Buss 7 (DB7)
22	Port C - 0	Output	Inductor Motor Driver STEP Pin	Inductor Motor Step
23	Port C - 1	Output	Inductor Motor Driver ENABLE Pin	Inductor Motor Power Enable
24	Port C - 2	Output	Switch Motor Driver DIR Pin	Switch Motor Direction Select
25	Port C - 3	Output	Switch Motor Driver STEP Pin	Switch Motor Step
26	Port C - 4	Output	Switch Motor Driver ENABLE Pin	Switch Motor Power Enable
27	Port C - 5	Output	Capacitor Motor Driver DIR Pin	Capacitor Motor Direction Select
28	Port C - 6	Output	Capacitor Motor Driver STEP Pin	Capacitor Motor Step
29	Port C - 7	Output	Capacitor Motor Driver ENABLE Pin	Capacitor Motor Power Enable
30	AVCC	Input	+5V DC Supply	ADC Power Supply
31	GND	-	Ground	
32	AREF	-	100 nF Capacitor - Internal Reference Selected	ADC 2.56 V Bandgap Voltage Reference
33	Port A - 7	-	Not Connected	
34	Port A - 6	-	Not Connected	
35	Port A - 5	-	Not Connected	
36	Port A - 4	-	Not Connected	
37	Port A - 3	-	Not Connected	
38	Port A - 2	-	Not Connected	
39	Port A - 1	Input (ADC1)	Reflected RF Power Sensor Voltage	ADC input from Reflected Pwr Sensor
40	Port A - 0	Input (ADC0)	Forward RF Power Sensor Voltage	ADC input from Forward Pwr Sensor

Table 7 - ATmega32 Microcontroller Pin Connections

Front Panel

The front panel LCD and buttons are connected to the microcontroller module via a 25 conductor cable. At the front panel, the cable wires are soldered to the buttons and LCD pins. The other end of the cable has a male DB25 connecting to the microcontroller module female DB25 allowing quick module removal. The LCD backlight requires 4.2-4.6 V DC power. To get this voltage, a 1N4001 diode was soldered in series with +5V DC wire and the backlight power pin. Assuming a minimum 0.5 V diode voltage, 4.5 V is supplied to the backlight. Figure 20 shows the complete IMS front panel.

Figure 20 - IMS Front Panel

Rear Panel

Figure 21 shows the IMS rear panel.

Figure 21 - IMS Rear Panel

Ceramic insulators isolate the balanced antenna transmission line terminals from the panel. A ground terminal provides an earth ground connection point for the chassis. The ground and antenna terminals use wing nuts allowing connection without tools. The IMS connects to a transmitter or transceiver via an SO-239 connector. The rear panel also contains the IMS power switch, power supply fuse holder, and 120 V AC connector.

Complete IMS

Figure 22 shows the IMS interior with all hardware construction complete. All internal RF connections on the source side of the matching network were made using RG-58 50 Ω coaxial cable. Copper tape covers the LCD backside for RF shielding.

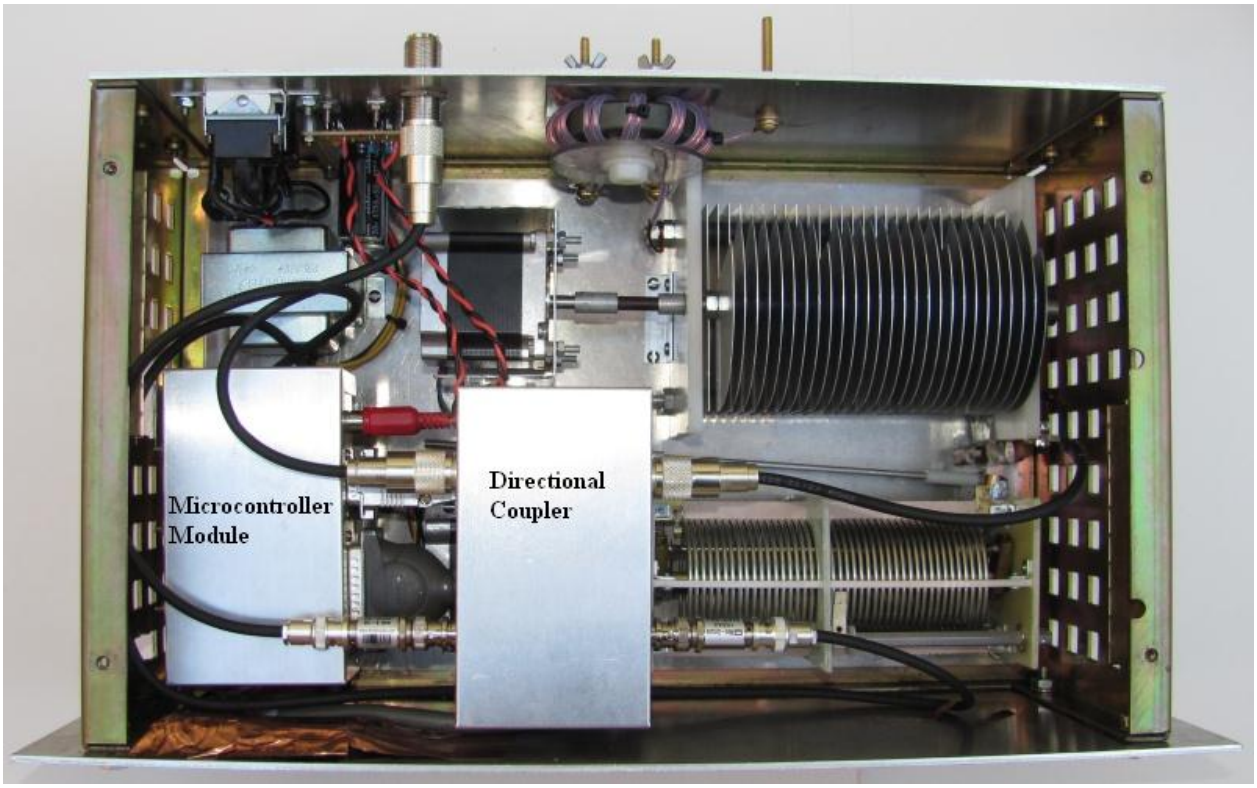

Figure 22 - IMS Interior

Software

The software was written after all hardware construction was complete. The completed software is shown in Appendix E. One feature at a time was enabled in the following order:

1. LCD Status Display
2. Manual Stepper Motor Control via Front Panel Buttons
3. Forward and Reflected Power Measurement and SWR Calculation
4. Automatic Impedance Matching

A more detailed software development discussion for these four features follows.

LCD Status Display

The LCD displays two lines of data with 16 characters each. Excel was used for display design. Figure 23 shows the Excel sheet.

1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16
F	_	_	_		R	_	_		S	W	R	_	_	.	_
	L	_	_	_	_		C	_	_	_		S	C	S	L

1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16
F	_	_	_		R	_	_		S	W	R	_	.	_	
>	L	_	_	_	_		C	_	_	_		S	L	S	C

Figure 23 - Excel Sheet LCD Display Configuration

A software function named “Update_Display” was written to write data to the display. LCD display software functions previously developed in the CPE-329 course were incorporated into the program.

Figure 24 shows the actual display.

Figure 24 - LCD Display

On the display, the number following “F” is the measured forward power in W. The number following R is the reflected power in W. SWR is calculated from the forward and reflected power and the result is displayed after the letters “SWR.” The number of motor steps counting up from minimum inductance and capacitance values are displayed after “L” and “C,” respectively. One of two possible RF switch positions, SCSL and SLSC, is displayed in the

lower right. The two possible positions are Series C-Shunt L, and Shunt L-Series C. This refers to the shunt inductor (L) being either on the source or load side of the series capacitor.

Manual Stepper Motor Control via Front Panel Buttons

A function named “check_buttons” was written to allow user manual stepper motor control. This allows user selection and actuation of individual stepper motors for the adjustable inductor or capacitor, or the RF switch.

The front panel “Select” button selects which motor to control. The “greater than” symbol (“ > ”) appears on the LCD indicating the selected motor. If the user presses the select button once, the “ > “ symbol appears in front of “ L” indicating that the inductor motor is selected. Pressing “Select” again places “ > “in front of “C” which indicates capacitor motor selection. Pressing “Select” a third time moves “ >” in front of the “SLSC”/”SCSL” switch position indicating that the switch motor is selected. Pressing “Select” a fourth time removes “ > “from the display indicating that no motor is selected.

When each motor is selected, the ENABLE line to the respective motor driver transitions from logic high (5 V) to low (0 V) causing the motor driver to apply power to the motor. With motor power applied, 12 V DC supply current is 1 A. To reduce power consumption, the software has a timer which de-selects a selected motor if no button has been pressed for ten seconds. With either the capacitor or inductor motor selected, the user presses the “Select +” button to increase capacitance or inductance. The “Select – “ button decreases inductance or capacitance. With the switch motor selected, pressing “Select +” toggles the switch position and the “SCSL”/ SLSC” display indication.

As the inductor or capacitor changes position, the displayed relative position values update. These values indicate the number of motor steps counting up from the minimum inductance and capacitance positions. The inductor value range is 0 to 8500 and the capacitor range is 0 to 1660.

Initial stepper motor testing determined 3200 steps per revolution rather than 200, the datasheet-specified number of steps per revolution. The motor driver instruction manual specifies that the drivers default to 1/16th step increments and that three pads on the driver require grounding to attain full steps. (16 * 200 steps/rev = 3200 steps/rev). Matching network testing determined optimum values of 200 steps/rev for the inductor and 3200 steps/rev for the capacitor.

Initially, the system did not recall motor position following a power cycle. The system must record position to avoid exceeding mechanical limits. Software functions were written to store the position values in the ATmega32 non-volatile EEPROM memory. EEPROM cells have a lifetime of 100,000 writes before failure; therefore the program must reduce EEPROM writing. The program writes values 10 seconds after the user has pressed any button and only if the values differ from those previously stored.

Forward and Reflected Power Measurement and SWR Calculation

A function named "Calculate_SWR" was written to obtain forward and reflected power values from the analog to digital converter (ADC) and convert them to power in W and SWR. Equation (14) [10] yields the RF power sensor output voltage for a given input power in dBm, P_{dbm}.

$$V_{out}, V = 0.025(P_{dBm}) + 2 \qquad (14)$$

For a 10 bit ADC, the number of possible output values is 2^{10} or 1024 numbers from 0-1023. Equation (15) gives the ADC decimal output value based on the ADC input voltage, V_{in}, and the ADC reference voltage, V_{REF}. [14]

$$ADC = 1024\left[\frac{V_{IN}}{V_{REF}}\right] \qquad (15)$$

The ADC input voltage is the power sensor output voltage from equation (14). The reference voltage, V_{REF}, was set to the ATmega32 internal 2.56 V bandgap voltage reference.
ADC V_{IN} = power sensor V_{out}.

A microcontroller limitation is the inability to perform floating point math; only integer numbers can be processed. To overcome this limitation, a lookup table (LUT) was created. Table 8 shows the spreadsheet used to calculate the LUT values.

IMS Look of Table Calculation					
	Vref, (V) =		2.560	Dir Coupler Factor (dB) =	50.00
	AD8307 Slope, (V/dB) =		0.025	AD8307 Y-intercept, (V) =	2.000
ADC	AD8307 Vout, (V)	AD8307 input power, dBm	Dir Coupler Input Power, dBm	Dir Coupler Input Power, W	LUT Value, 100*W
0	0.000	-80.0	-30.0	0.0000010	0
1	0.003	-79.9	-29.9	0.0000010	0
2	0.005	-79.8	-29.8	0.0000010	0
3	0.008	-79.7	-29.7	0.0000011	0
...	...	...	...	...	...
600	1.500	-20.0	30.0	1.0000000	100
...	...	...	...	...	...
1023	2.558	22.3	72.3	16982.4365246	1698244

Table 8 - IMS Power Look Up Table Calculation

The AD8307 V_{out} is calculated for each possible ADC value from 0-1023 using equation (16).

$$V_{out} = V_{REF}\left[\frac{ADC}{1024}\right] \qquad (16)$$

AD8307 input power, P_{IN}, in dBm, is calculated using equation (17).

$$P_{IN}, dBm = \frac{V_{out} - [AD8307\ Y - int, (V)]}{\left[AD8307\ Slope, (\frac{V}{dB})\right]} = \frac{V_{out} - 2V}{\left[0.025\ (\frac{V}{dB})\right]} \qquad (17)$$

The directional coupler input power is calculated by adding the 50 dB coupling factor to the AD8307 input power.

The directional coupler input power is converted to W using equation (7). The ATmega32 microcontroller cannot perform the equation (7) exponential function since it requires a floating point operation. This necessitates the LUT.

The LUT contains pre-calculated results of the equation (7), (16), and (17) math functions. The LUT shares 32 kB memory with the 8 kB software program limiting its size to 24 kB or 1536 values at 16 bits each. 16 bit values can represent integers up to 2^{16} or 65,536. Since the microcontroller cannot work with fractional numbers, they must be converted to whole integer numbers by multiplying up to larger value numbers, performing the required calculations, and then using integer division to obtain the final result. This method preserves some of the accuracy that would otherwise be lost if all numbers were rounded to whole numbers.

All digital systems have accuracy limitations since values between consecutive binary values cannot be defined. In contrast, analog systems can process an infinite number of values. For example, if a 10 bit digital system must represent a voltage range from 0 to 1 V, it can only represent 210 or 1024 possible values between 0 and 1 V. In an analog system, the voltage can assume an infinite number of possible values between 0 and 1 V. The designer must use a system with enough bits to ensure the required accuracy.

The IMS minimum power measurement required is 10 mW or 10 dBm. At the lowest IMS RF input power of 3 W, 10 mW reflected power gives 1.12 SWR using equation (2). If the minimum measurable power is limited to 10 mW, the IMS could attain a minimum SWR of 1.12, which is less than the IMS matching capability requirement 1.5 SWR or less.

Since the IMS maximum power requirement is 100 W, its power measurement ability must be greater than 100 W. A maximum of 200 W was arbitrarily chosen. Therefore, the total required measurement range is 10 mW – 200 W or 10 dBm to 53 dBm. From Table 8, the power values are in 0.1 dB steps. With a 43 dB measurement range, the required LUT size is 430 values. With a start value of 0.01 W, the LUT values must be multiplied by 100 to convert all values to integers. The system divides the values by 100 to get the power in W prior to displaying on the LCD.

From equation (14), -80 dBm corresponds to 0 V power sensor output. With the IMS power measurement range limited from 10 dBm to 53 dBm, the first LUT value corresponds to 10 dBm. From equations (16) and (17), the corresponding ADC value is 400. The LUT is written in a C programming structure called an array, which is a programmable number of memory locations storing the same data type under the same variable name. Array values are accessed by inserting a reference number in brackets following the array name. For the array named "Power_LUT", the program accesses the first value using "Power_LUT[0]." With the LUT size set to 430, the last value is accessed using "Power_LUT[429]." The IMS LUT table converts ADC values to power in W. The LUT begins at an ADC value of 400. The ADC results are stored in a variable named "power_index," which then has 400 subtracted from it. It is then used to select values in the LUT. ADC values below 400 are set to zero; therefore the IMS registers power below 10mW as 0W. The following code shows how the LUT power value is assigned in the C program. Comments in green explain each line of code.

```
Power_index = ADC;          // store ADC converted value to Power_index variable
   if (Power_index < 400)   // this makes any power less than offset = 0 W
       Power_index = 0;
   else
       Power_index = Power_index - 400;   // index of proper value in LUT
       Forward_Power = Power_LUT[Power_index]; //store power to Forward_Power variable
```

The IMS power meter can be calibrated in software. If the measured power is a constant dB value off from the actual power over the full measurement range, the 400 offset value can be changed to reference offset power values in the LUT. Since each consecutive LUT value is a 0.1 dB change, the dB adjustment value required for calibration multiplied by 10 is the required offset change. For example, if the indicated power is 50W when the actual power is 100W, a 3 dB difference, then the 400 offset must be changed by 30. Changing the offset number from 400 to 370 corrects the 3 dB error by referencing 3 dB higher values in the LUT.

If the power measurement error is not a consistent dB value, there is a slope error. For example, the power is 3 dB off at 10 W and 6 dB off at 100 W. This error requires recalculating the LUT using a different slope in equation (17).

Automatic Impedance Matching

To develop the IMS automatic impedance matching ability, three algorithms were written and tested

First algorithm:

1. For RF input power 5 W or greater, begin automatic tuning when Tune button is depressed
2. Measure and record current SWR
3. Rotate inductor one step in increasing direction
4. If SWR decreased, repeat 2.
5. If SWR increased, rotate inductor one step in decreasing direction
6. If SWR decreased, further decrease inductance

7. When SWR minimum is determined, repeat 1 through 5 for the capacitor.
8. If SWR is greater than 1.5:1, repeat for the inductor and then capacitor up to 3 times.
9. Stop automatic tuning attempt if the Tune Button is pressed again or if RF input power falls below 5 W.

This first algorithm adjusts inductance and capacitance based on SWR increase or decrease. This approach assumes that inductor or capacitor value variations in one direction consistently increase or decrease the SWR until the minimum is reached. This may hold true for an average over several SWR samples. In actuality, due to measurement noise or random fluctuation, the SWR may increase for small adjustments even if the adjustments are in the correct direction to achieve an impedance match.

A second algorithm was written using a different approach as follows:

1. Measure and store current SWR as minimum SWR.
2. Set the inductor motor direction to increase inductance.
3. Change inductance by one motor step
4. Measure SWR.
5. If the new SWR is less than the stored minimum value, store the new value as the minimum SWR and store the inductor position.
6. If the SWR increased, increment a counter tracking the number of times the SWR increased.
7. Repeat Steps 3 through 6 until the number of steps for which the SWR increased exceeds a set value (increasing SWR step count, optimum value determined through trial and error). Change the motor direction to decrease inductance.
8. Repeat Steps 3 through 6 after the Step 7 direction change until the increment counter exceeds the set value (increasing SWR step count, optimum value determined through trial and error).
9. Return the inductor to the minimum SWR position.
10. Repeat Steps 1 through 9 for the capacitor.
11. Repeat Steps 1 through 10 up to 4 times or until the SWR decreases below 1.5.
12. If the SWR is still greater than 1.5, stop the tuning process.

This algorithm failed as it would only attain an impedance match if the required inductor and capacitor positions were within the set number of increasing SWR steps. Setting the increasing SWR step count to higher values, from 20 to 200, improved match success rate. However, the full range is 1660 steps for capacitance and 3500 steps for inductance. These full ranges must be searched to find the optimum match values. The inductor has a full range of 8500 steps, however for all loads tested, the inductor never required a setting beyond 3500. Therefore, the inductance search range was set to 0-3500.

A third and final algorithm was written:

1. Set the inductor to minimum inductance position and the capacitor to maximum capacitance position.

2. Measure and store current SWR as minimum SWR.
3. Increment the inductor from minimum to maximum value. For each motor step:
    a. Measure SWR.
    b. If the new SWR is less than the stored minimum value, store the new value as the minimum SWR and store the inductor position.
4. Once the maximum inductance is reached, return the inductor to the location of minimum SWR.
5. Repeat steps 2 through 4 for the capacitor, stepping from maximum to minimum capacitance.
6. If the SWR is still greater than 1.5, perform Steps 7 through 17.
7. Measure and store current SWR as minimum SWR.
8. Set the motor direction to increase inductance.
9. Change inductance by one motor increment.
10. Measure SWR.
11. If the new SWR is less than the stored minimum value, store the new value as the minimum SWR and store the inductor position.
12. If the SWR increased, increment a counter tracking the number of times the SWR increased.
13. Repeat Steps 9 through 12 until the number of increments for which the SWR increased exceeds a set value. Change the motor direction to decrease inductance.
14. Repeat Steps 3 through 6 after the direction change in Step 13 until the increasing SWR increment counter exceeds the set value (increasing SWR step count, optimum value determined through trial and error).
15. Return the inductor to the minimum SWR position.
16. Repeat Steps 7 through 15 for the capacitor.
17. Repeat Steps 7 through 16 up to four times or until the SWR decreases below 1.5.
18. If SWR is not below 1.5, stop the tuning process.

Steps 1 through 5 perform low resolution (coarse) tuning and Steps 7 through 17 perform fine tuning.

Optimum settings determined through trial and error:

Inductor step angle: 1.8°/step, or 200 steps/revolution
Capacitor step angle: 0.1125°/step, or 3200 steps/ revolution
Inductor increasing SWR step count: 110
Capacitor increasing SWR step count: 250
Inductor coarse tuning step time interval: 2 ms
Capacitor coarse tuning step time interval: 10 ms
Inductor fine tuning step time interval: 13 ms
Capacitor fine tuning step time interval: 10 ms

Chapter 7. Integration and Test Results

1. Directional Coupler

The following equipment was used for directional coupler testing:

- Vector Network Analyzer, HP8753A
- Spectrum Analyzer, Agilent N9000A
- RF Signal Generator, Fluke 6060B
- Attenuator, Coaxial 20±0.2 dB fixed, 50 Ω, 0.5 W, DC – 2 GHz, Mini-Circuits HAT-20+, Quantity: 2
- 50 Ω termination, Anritsu

Test 1: RF Input Port S11

The first test performed was input port S11 measurement to ensure that the directional coupler would not present high standing wave ratio (SWR) values to a connected transmitter. The directional coupler input SWR was measured using the Figure 25 setup.

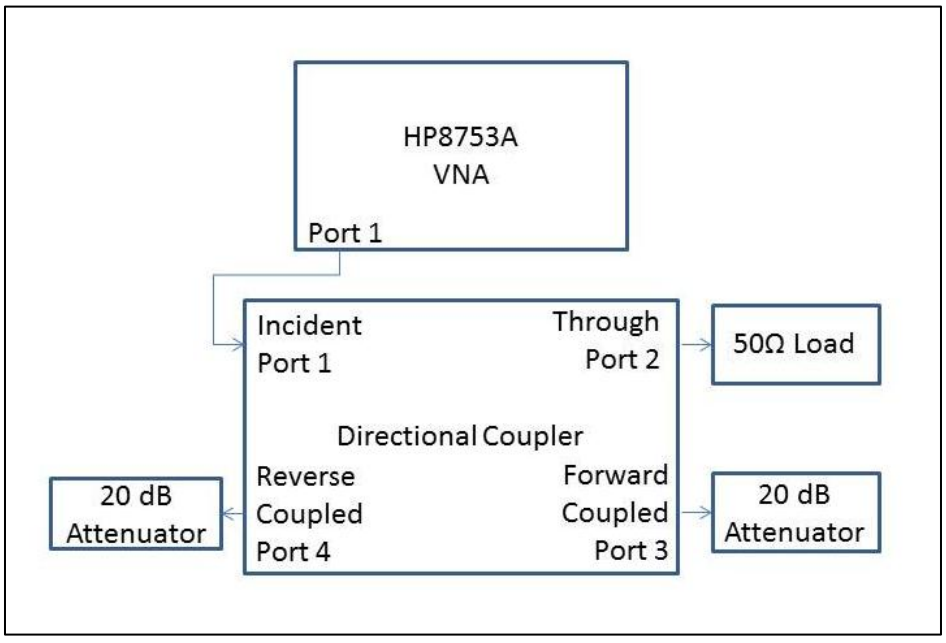

Figure 25 - Directional Coupler SWR Test Setup

The 20 dB attenuators served as 50 Ω terminations.

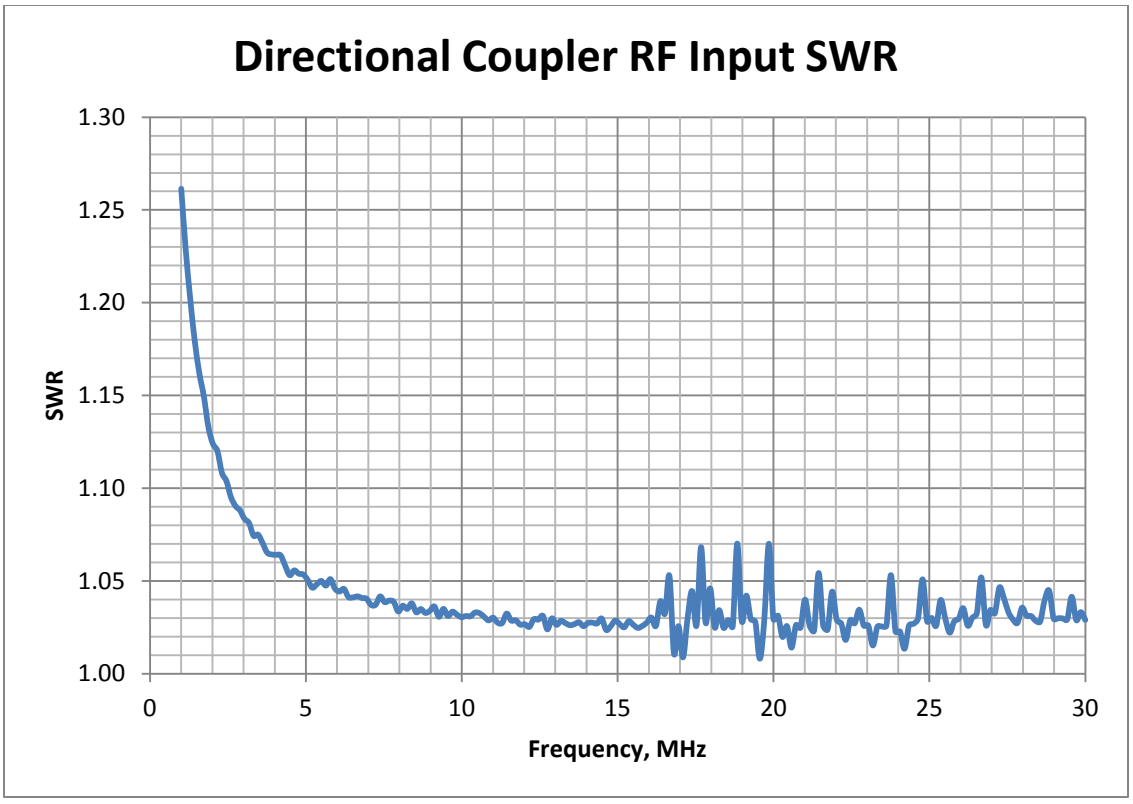

Figure 26 - Directional Coupler RF Input Port SWR

Figure 26 shows the RF input port SWR measured using the VNA. The worst case SWR for 3.5-30 MHz is 1.07 at 3.5 MHz. The directional coupler SWR must be less than 1.5 for the IMS to meet its matching requirement of 1.5 SWR or less.

Test 2: Coupling Factor

The forward coupling factor is the dB difference between the power measured (dBm) at the incident (RF input) port and forward coupled port. This was measured using the Figure 27 setup.

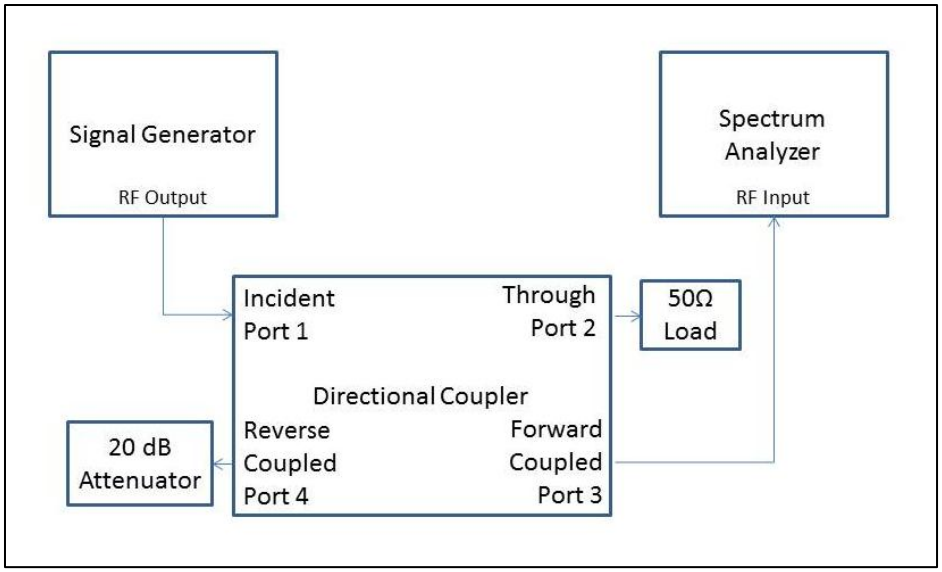

Figure 27 - Forward Coupling Factor Test Setup

The spectrum analyzer was set for 1 – 31 MHz span and placed in "max hold" mode. The signal generator was set to 0 dBm and the frequency was swept from 1 – 31 MHz. The spectrum analyzer "max hold" mode captured the power measured from the forward coupled port. With the signal generator set to 0 dBm, equation (18) yields coupling factor.

$$coupling\ factor, (dB) = 0\ dBm - measured\ power, (\ dBm) \qquad (18)$$

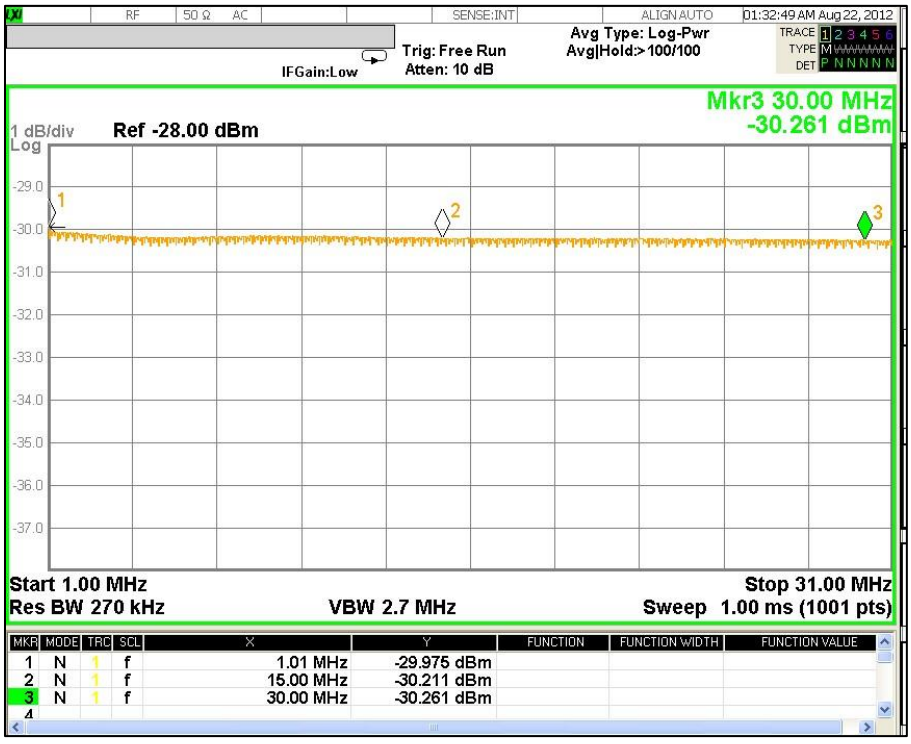

Figure 28 - Directional Coupler Forward Coupling Factor

Figure 28 shows the spectrum analyzer forward coupling factor measurement. The forward coupling factor is approximately 30 dB and varies by less than 0.5 dB from 1 – 31 MHz. This is the expected result per Kaune [11].

The reflected coupling factor is the dB difference between through port and reverse (reflected) coupled port power. This was measured using the setup shown in Figure 29 using the same procedure for the forward coupling factor measurement.

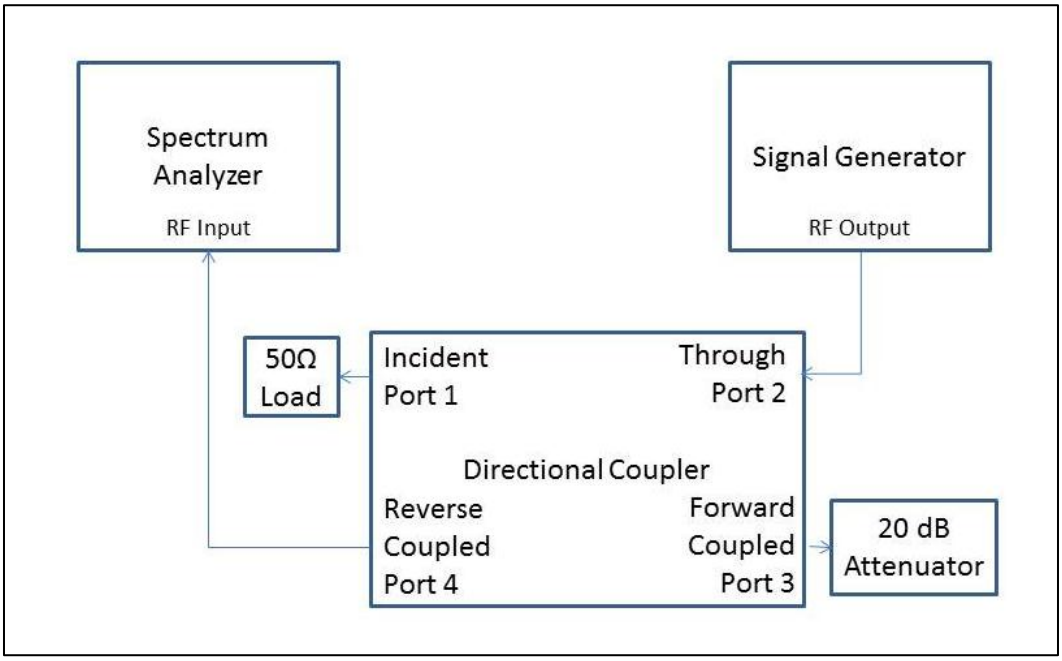

Figure 29 - Reverse Coupling Factor Test Setup

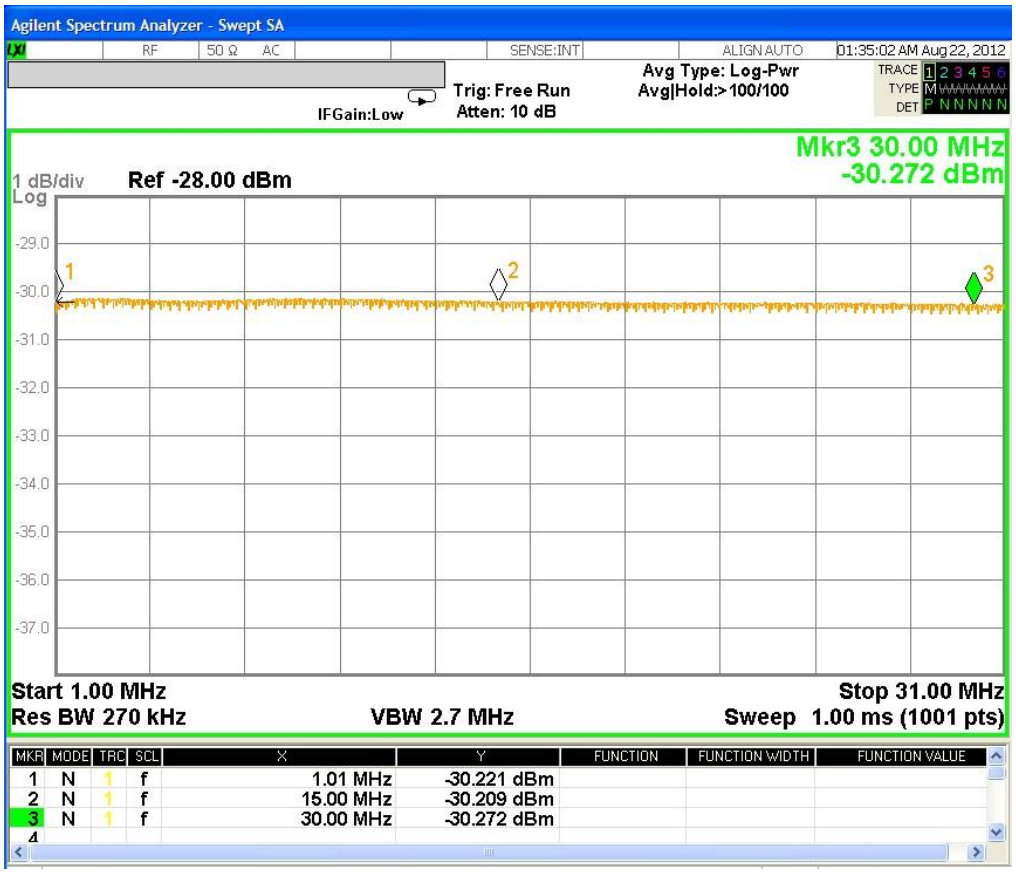

Figure 30 - Directional Coupler Reverse Coupling Factor

Figure 30 shows the spectrum analyzer reverse coupling factor measurement. The reverse coupling factor is approximately 30 dB and varies by less than 0.5 dB from 1 – 31 MHz. This is the expected result per Kaune [11].

Test 3 – Directivity

The directional coupler directivity is the dB difference between forward and reflected coupled powers with power applied at the incident port and the through port terminated in 50 Ω. The reverse coupled port power with power applied at the incident port was measured using the Figure 31 setup. The signal generator was again set to 0 dBm and swept from 1 – 31 MHz.

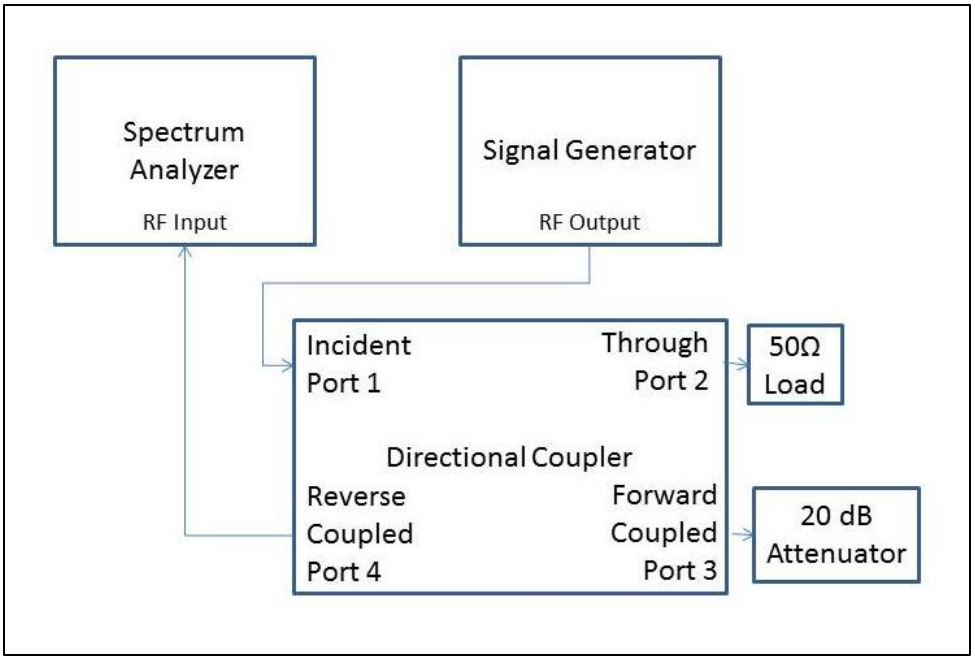

Figure 31 - Directional Coupler Directivity Test Setup

The directivity was calculated by subtracting the Figure 31reverse port power measurement from the forward port power measurement. The resulting directivity is plotted in Figure 32.

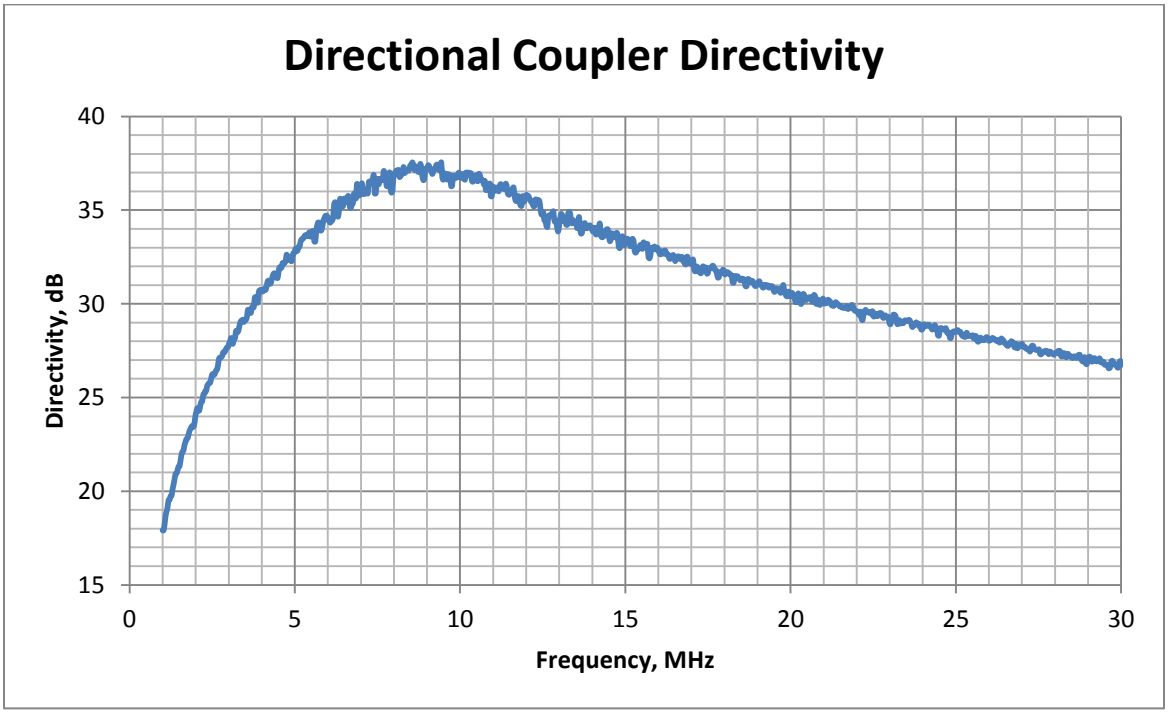

Figure 32 - Directional Coupler Directivity

The worst case directivity for 3.5 – 30 MHz is 26.56 dB at 29.65 MHz. Kaune claimed a worst case directivity of 28 dB [11]. Directivity is ideally infinite.

For 100 W (50 dBm) forward power and 30 dB coupling factor, the forward coupled power is 20 dBm or 0.1 W. With 28 dB of directivity, the reflected coupled port power is:

$$20 \text{ dBm} - 26.56 \text{ dB} = -6.56 \text{ dBm} = 0.22 \text{ mW} \qquad (19)$$

Using equation (2) to calculate SWR from forward and reflected coupled power, the SWR measured using this directional coupler is 1.1.

The IMS power measurement limit is determined as follows:

Given:
Directional coupler coupling factor = 30 dB
Maximum RF power sensor input power = 17 dBm
Maximum Directional Coupler Input Power = 17 dBm + 30 dB = 47 dBm (50 W) (20)

Add 20 dB , 0.5 W (27 dBm), attenuators at forward and reflected coupled ports:
Maximum Directional Coupler Input Power = 27 dBm + 30 dB = 57 dBm (500 W) (21)
RF sensor input power at 500W = 57 dBm – 20 dB – 30 dB = 7 dBm (22)

The IMS maximum RF power requirement is 100 W.

2. RF Power Sensor

The forward and reflected RF power sensors were tested for DC voltage output linearly proportional to RF power input using the test setup shown in Figure 33.

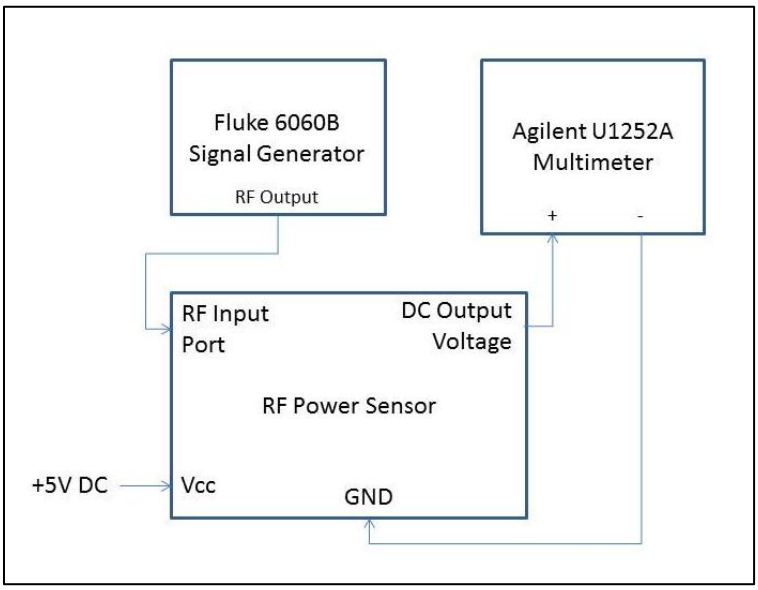

Figure 33 - RF Power Sensor Test Setup

The AD8307 datasheet [10] specifies a measurement capability of -75 dBm to +17 dBm. The sensors were tested in this range, but the voltage output was non-linear below -50 dBm as shown in Figure 34. The Fluke 6060B signal generator maximum output power is 13 dBm.

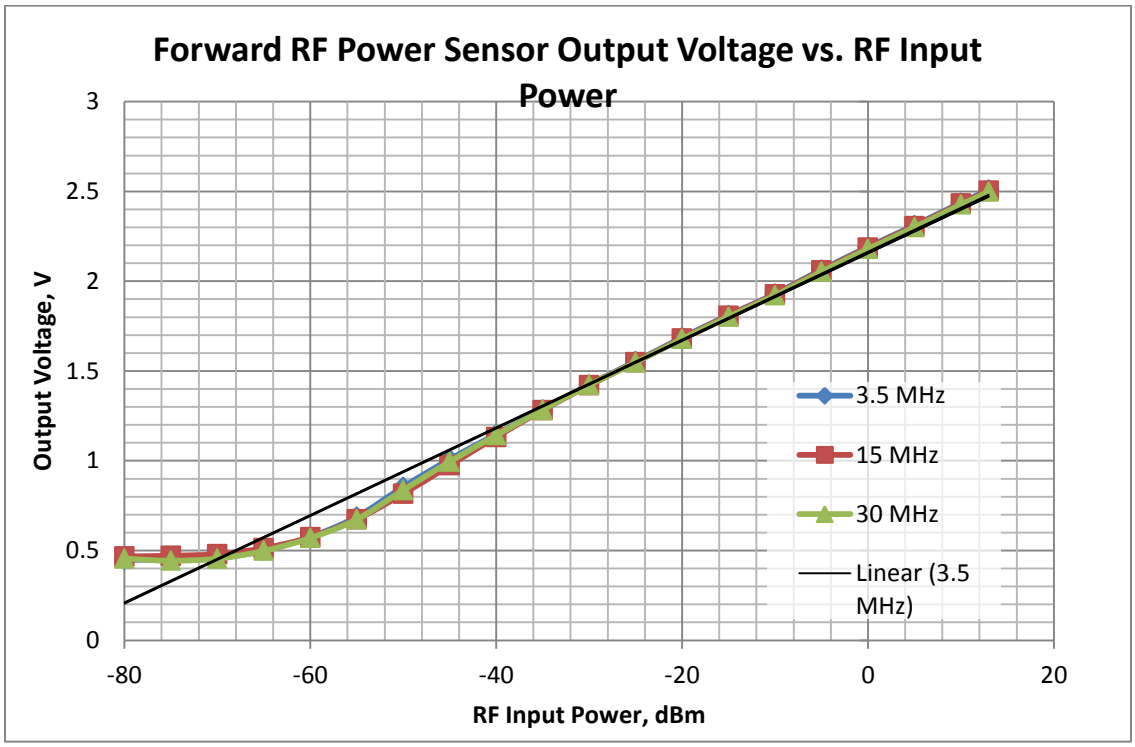

Figure 34 - Forward RF Power Sensor Output Voltage vs. Input Power, -80 to 13 dBm

The IMS minimum power measurement is 10 mW. At minimum IMS RF input power of 3 W, 10 mW reflected power yields 1.12 SWR using equation (2). If the minimum measurable power is limited to 10 mW, the IMS could calculate a minimum SWR of 1.12, which is less than the IMS matching requirement of 1.5 SWR or less.

For 10 mW (10 dBm) reflected power, RF power sensor input is -40 dBm after 30 dB directional coupler coupling factor and 20 dB attenuator. Therefore, the minimum RF sensor measurement is -40 dBm.

The power sensors were tested from -40 dBm to +13 dBm at 3.5, 15, and 30 MHz. Figure 35 and Figure 36 show forward and reflected power sensor output voltage responses, respectively.

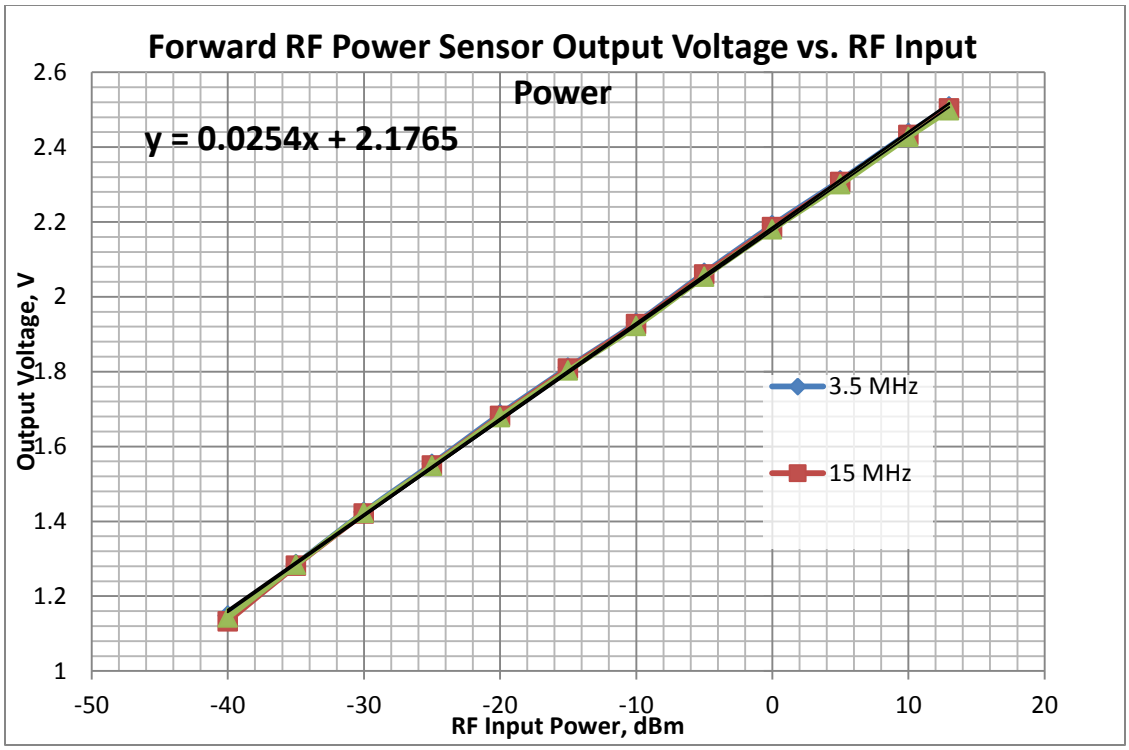

Figure 35 - Forward RF Power Sensor Output Voltage vs. Input Power, -40 to 13 dBm

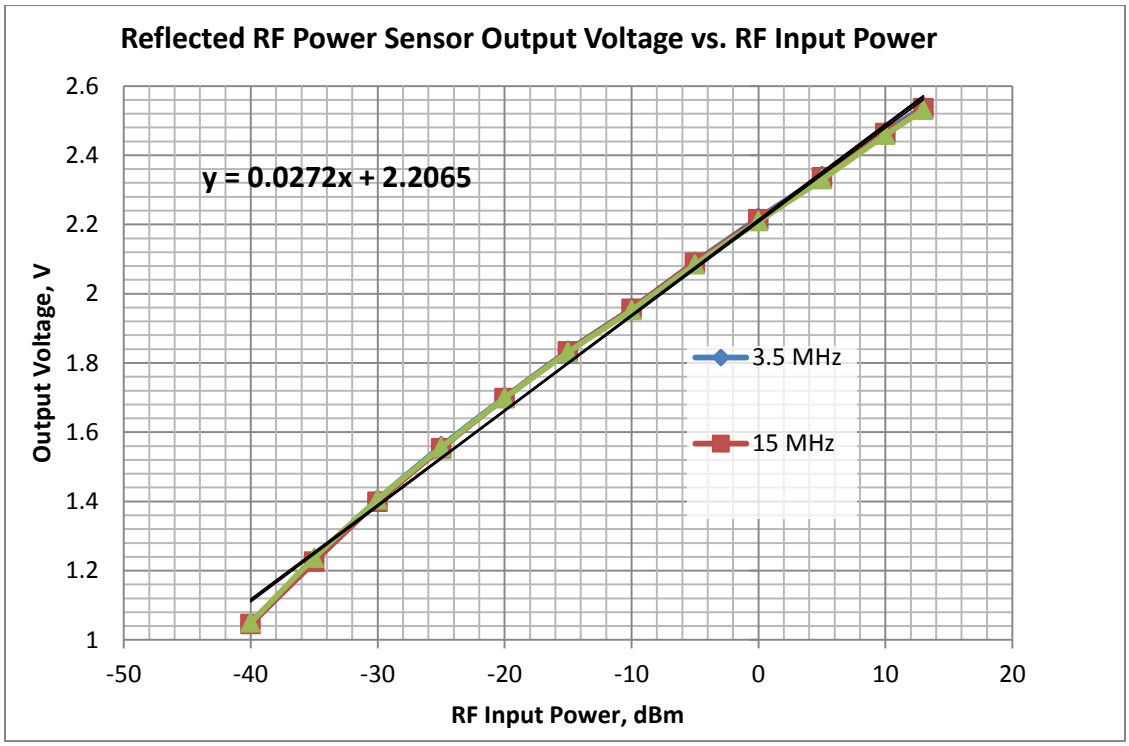

Figure 36 - Reflected RF Power Sensor Output Voltage vs. Input Power, -40 to 13 dBm

The curve-fit equations shown in Figure 35 and Figure 36 are linear fit relations for the data. The equation slope is the AD8307 datasheet parameter "Logarithmic Slope" in units of V/dB. The measured forward and reflected slopes are 25.4 mV/dB and 27.2 mV/dB, respectively. The datasheet specifies 23 – 27 mV/dB from -75 dBm to + 17 dBm. The equation Y-intercept should be 1.9 – 2.2 V [10].

3. Power Supply

Table 9 shows the power supply voltages tested as specified in the Chapter 6 Test Plan.

DC Power Supply Test Data				
	Output Voltage, V			
	No Load	Full Load	Design Requirement	Datasheet Specification [16] [17]
5 V, 0.5 A DC Supply	5.031	5.028	4.7 to 5.5	5±0.05
12 V, 1 A DC Supply	12. 07	11.94	11 to 13	12±0.6

Table 9 - DC Power Supply Test Data

The supply voltages were measured at the microcontroller module power connectors with applied system power. With all stepper motors off, the 12 V and 5 V supplies measured 12.070 V and 5.028 V, respectively. With the inductor motor on, the 12 V and 5 V supplies measured 11.940 V and 5.028 V, respectively. These voltages are still within the datasheet specifications for the 5 V and 12 V regulators [16] [17].

4. Stepper Motors

Figure 37 shows the stepper motor test setup. This test determined whether the motor produces the torque required for variable inductor and capacitor rotation. Since the motor successfully rotated the capacitor and inductor, the required torque was less than the maximum rated motor torque of 125 oz-inch. A method to measure the actual required torque was not available.

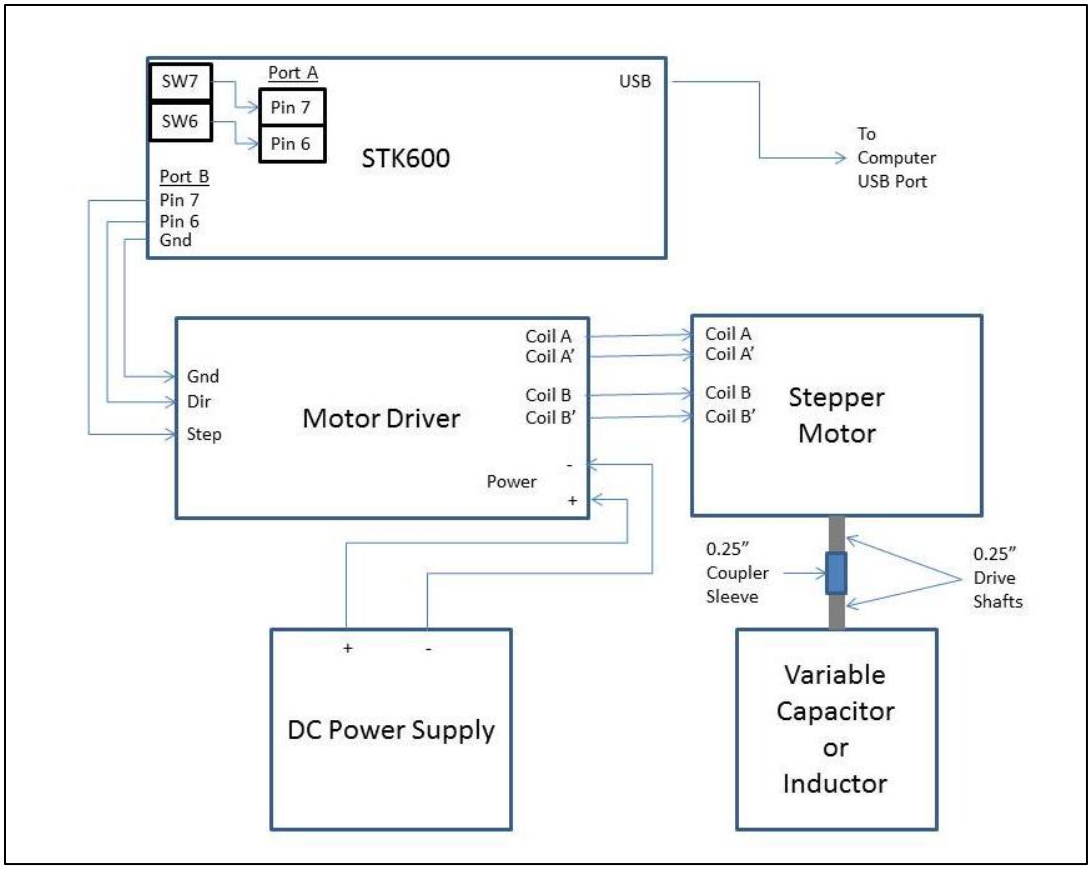

Figure 37 - Stepper Motor Test Setup Block Diagram

The STK600 Development board provides power, clock signal, and programming interfaces to the microcontroller. It also provides large pin breakouts for all input and output ports allowing solderless connections.

The stepper motor driver provides the required power waveforms to the stepper motor. The two motor coils, A and B, must be driven by 90° out of phase square waves. The motor drive waveforms rotate the motor one step for each rising edge, 0 V to 5 V, of a digital signal applied to the STEP input. This simplifies the microcontroller program used to control the stepper motor. The DIR input controls rotational direction; high (5 V), counterclockwise, low (0 V), clockwise. [20]

Stepper motor specifications give motor current for a specified voltage. The IMS motors are rated for 2 A 3 V DC [18]. The motor can function with higher voltage at a proportionally lower current (V*I = 6 W maximum). For the test, the voltage was set to 12 V, requiring 0.5 A.

When power is applied to the driver, the motor draws the same current regardless of whether it is moving or stationary. This maintains mechanical loads that rotate away from the required position. Since the variable capacitors and inductors remain stationary due to static friction, the motors can be powered down when no adjustments are required, saving power. The initial lack of clarity on this stepper motor aspect caused some confusion as to whether there was a driver malfunction or incorrect driver to motor wiring.

Figure 38 shows the capacitor stepper motor test setup. The motor was mounted using a makeshift setup for testing purposes. A permanent mounting fixture was fabricated for installation in the antenna tuner. The capacitor's full value range is spanned in a half revolution. At the top motor speed tested, the capacitor is actuated through its full range in less than one second.

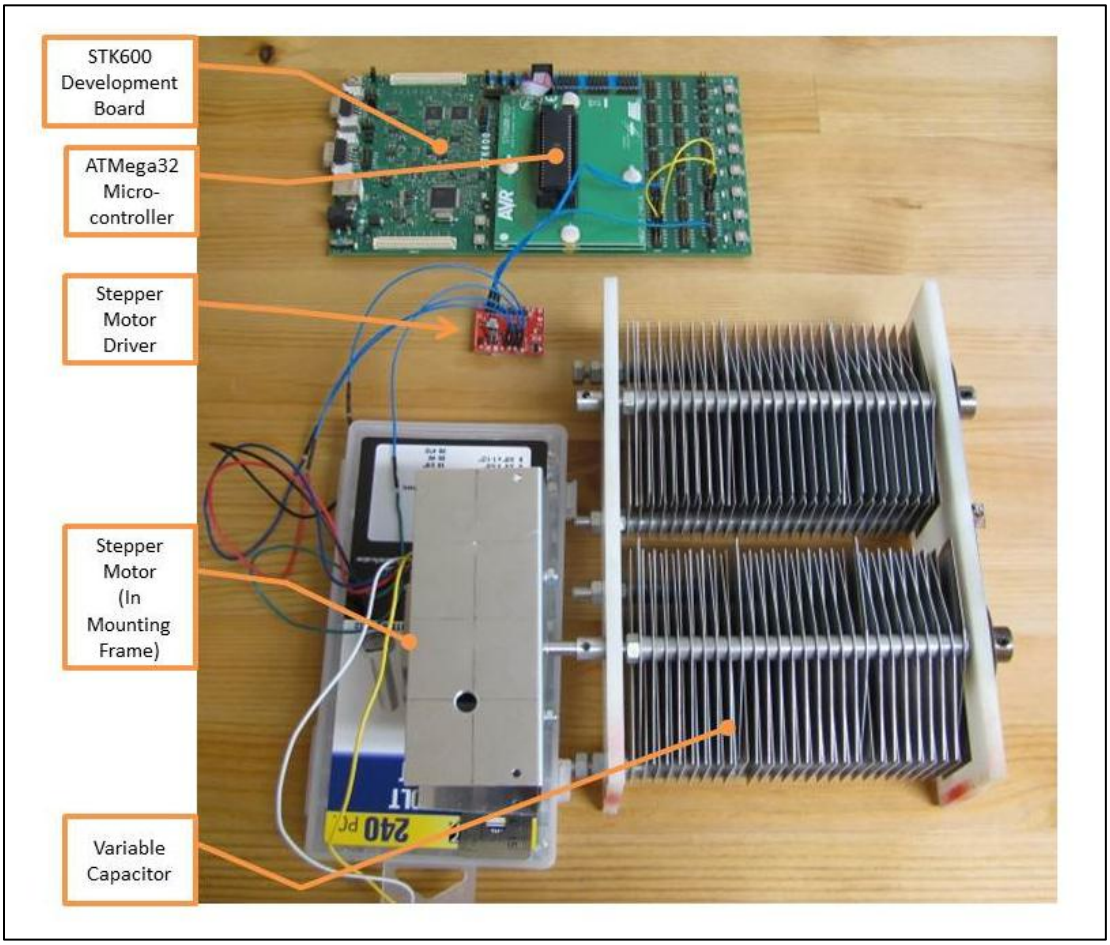

Figure 38 - Capacitor Stepper Motor Test Setup

In Figure 38, the motor driver DC power supply leads and the STK600 USB cable are not yet connected. Two variable capacitors are attached to a common frame. Only one is required for the IMS.

The stepper motor is unipolar; each coil has three leads: two outer leads and a center tap. The motor is unipolar because the power supply current does not change polarity. Positive supply voltage is connected to the center tap and the driver grounds either outer lead depending on the required winding field direction. [23] Bipolar motors have two leads for each coil. The driver reverses coil voltage polarity to change the field direction. The motor driver, while designed for bipolar motors, can drive unipolar motors if the center tap leads are not connected. [20] Figure 39 shows the unconnected white and yellow center tap motor leads.

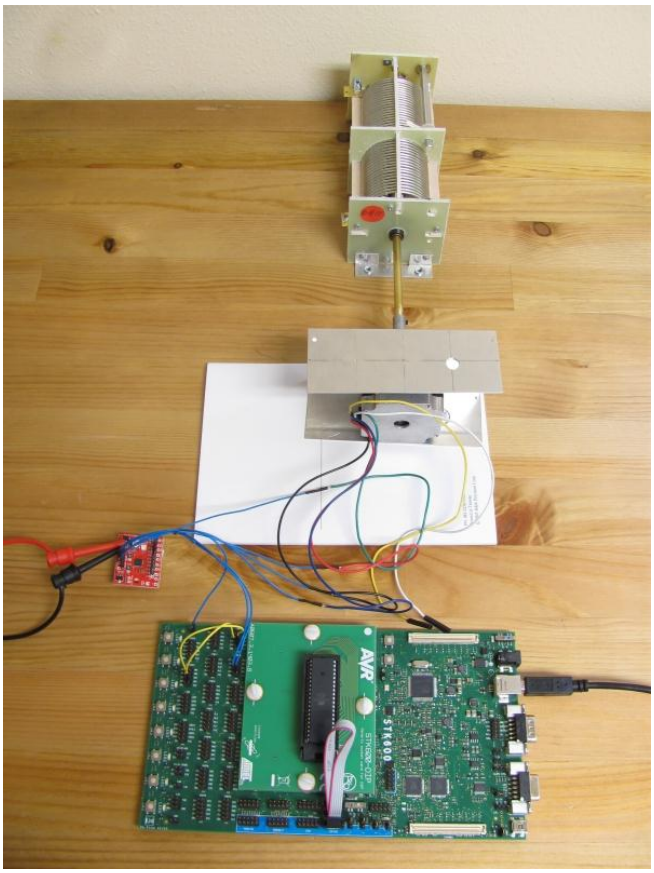

Figure 39 – Variable Inductor Motor Test Setup

Figure 39 shows the variable inductor drive test configuration. At the maximum motor speed setting, the inductor travels through its full range in approximately 20 seconds. Full range traversal requires 43 shaft revolutions.

```
1  //-------------------------------------------------------------
2  //- description: this program is used to test a stepper motor
3  //- using the Atmel ATmega32 microcontroller on an
4  //- STK600 development board and the "Big Easy Driver" stepper
5  //- motor driver. Port A connects to the buttons on the STK600 and
6  //- Port B connects to the motor driver.
7  //- -------------------------------------------------------------
8
9  #include <avr/io.h>          // AVR Standard Input/Output Library
10 #include <util/delay.h>      // Delay Function Library
11
12 #define F_CPU 3683400UL      // clock frequency for delay functions
13
14 int main(void)
15 {
16    DDRA = 0x00;      //Set port A input for buttons
17    DDRB = 0xFF;      // set PORT B as output for motor driver
18    PORTB = 0x00;     // initialize Port B to all zeros
19
20    while (1)         // infinite loop to keep program running
21    {
22
23         // This IF statement pulses the Step bit to drive the motor
24         // continuously as long as button 7 is pressed
25         if (!((PINA) & (0x80)))      //if button 7 pressed
26             {
27                 _delay_us(100);      // delay to control motor speed
28                 PORTB |= (1 << 7);   // make Step go high
29                 _delay_us(100);      // delay to control motor speed
30                 PORTB &= ~(1 << 7); // make Step go low
31                 _delay_us(100);      // delay to control motor speed
32             }
33
34         // This IF statement toggles the DIR bit for button 6 press
35         // to change the motor direction clockwise or counter-clockwise
36         if (!((PINA) & (0x40)))      //if button 6 pressed
37             {
38                 _delay_ms(200);      //delay for button debounce
39                 PORTB ^= (1 << 6);   //toggle DIR bit
40                 _delay_ms(200);      //delay for button debounce
41             }
42    }
43    return (0);
44 }
45
```

Figure 40 – C program for stepper motor testing

Figure 40 shows the C program written for microcontroller motor testing. Comments in blue text explain the function of each line. Basic functions include motor rotation using one button (or momentary switch) and rotation direction selection using a second button. Two IF statements accomplish these two functions. The first IF statement steps the motor continuously when Button 7 is pressed. The "_delay_us(x)" function defines a software delay of x microseconds. The delays in the first IF statement set the motor speed. Increasing the delay decreases motor speed. The second IF statement changes the motor direction each time Button 6 is depressed. The button numbers are labeled on the STK600.

Since the motor can drive the capacitor and inductor directly, gear reduction is not required. The inductor drive speed rotates the antenna tuner at slower rates than relay selected fixed inductor designs. Turning the inductor faster may increase wear, decreasing component life.

5. IMS Power Meter

The Impedance Matching System (IMS) requires power measurements and SWR calculation to achieve an impedance match. Uncalibrated, the IMS indicated 2:1 SWR for a matched load.

Calibration requires comparison to measurement devices of known accuracy. The most accurate power meter available is the Agilent N9000A Signal Analyzer, with specified amplitude accuracy of ±0.5 dB [24]. Its +30 dBm maximum input power limits usefulness for calibrating a 100 W (50 dBm) power meter. A 30 W, 30 dB attenuator allows power measurements to +44.8 dBm (30 W). Since power levels greater than 30 W could damage the attenuator and the N9000A, a different power meter was required for 30 W to 100 W calibration.

Three wattmeters with 100 W ratings or greater are available: A Bird 43 analog wattmeter with a 2 MHz -30 MHz, 250 W element; a Motorola R22 0.1 to 1000 MHz Communications Service Monitor (measures up to 100 W); and a Daiwa CN-801 1.8 to 200 MHz, 2000 W analog wattmeter. These wattmeters were compared to the Agilent N9000A spectrum analyzer in power measurements up to 30 W in Table 10. The 30 dB attenuator was connected to the N9000A RF input for protection.

Transmitter Power Setting (0 to 10 scale)	Agilent N9000A Spec. An., W	N9000 A Spec. An. , dBm	Bird Wattmeter, W	Bird, dBm	Bird & Spec. An, dB difference	Motorola R22 Communications Service Monitor, W	Motorola R22, dBm	Motorola & Spec. An. Db difference	Daiwa Wattmeter, W	Daiwa, dBm	Daiwa & Spec. An. dB diff
0	3.5	35.5	3	34.8	-0.7	3.2	35.1	-0.4	2.8	34.5	-1.0
1	5.6	37.5	5	37.0	-0.5	6.4	38.1	0.6	4	36.0	-1.5
2	9.5	39.8	10	40.0	0.2	11	40.4	0.6	7.2	38.6	-1.2
3	17.0	42.3	14	41.5	-0.8	14.9	41.7	-0.6	14	41.5	-0.8
4	25.7	44.1	25	44.0	-0.1	20.4	43.1	-1.0	21	43.2	-0.9

Table 10 - Power Meter Comparison

The Bird wattmeter was the most accurate with a maximum 0.8 dB difference from the N9000A. This unit was used for power calibration from 30 W to 100 W.

The IMS power meter was calibrated using the setup shown in Figure 41 for 0 W to 30 W calibration and Figure 42 for 30 W to 100 W calibration. To calibrate the reflected power measurement, the directional coupler RF input and output ports were interchanged; RF power was applied to the output port.

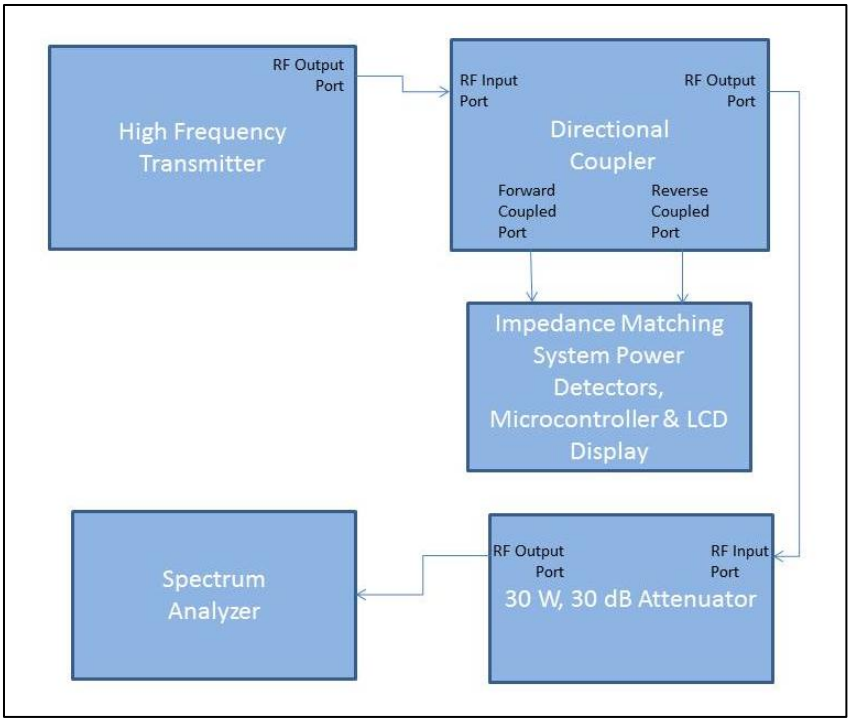

Figure 41 - IMS 0 W to 30 W Power Calibration Setup

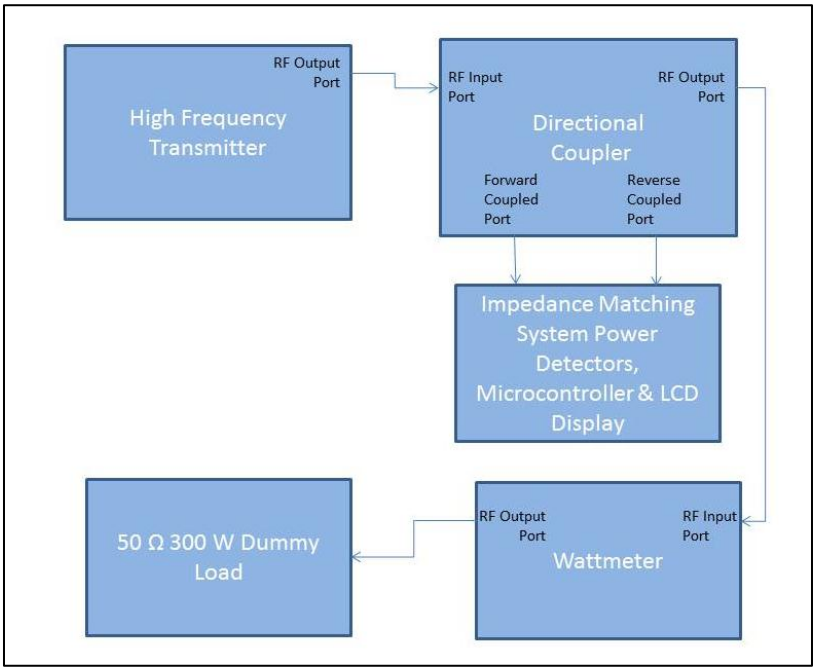

Figure 42 - 30 W to 100 W Power Calibration Setup using Bird 43 Wattmeter

Initially, the IMS forward power indication averaged 2.5 dB less than spectrum analyzer measurements. Since each consecutive value in the IMS LUT represents a 0.1 dB power increment, an offset number of 25 (25 * 0.1 dB = 2.5 dB) was added to the "power_index" variable in the C program as described in Chapter 6. The IMS reflected power was an average of 3.6 dB less than SA values; an offset of +36 was used.

After comparing the IMS power meter to the SA, it was compared to the Bird wattmeter. The Bird wattmeter scale increments are 5 W for the 250 W element. Estimating the halfway point between increments allows ±2.5 W measurement certainty. Therefore, the IMS power meter displays values within ±2.5 W of the Bird wattmeter.

Table 11 and Table 12 show power measured by the Agilent N9000A spectrum analyzer (SA) compared to IMS measured power after calibration. The SA measured power was adjusted by 30 dB to account for the 30 dB attenuator. Since the IMS can only display whole number power values, due to LCD character limit, the values in the IMS measured power column are shown to zero decimal places. As Tables 10 and 11 show, the IMS power is a maximum of 0.66 dB off from the SA measurement.

Forward Power					
Transmitter Pwr Setting (0 to 10 scale)	Spec An Pwr, dBm	Spec An Pwr, W	IMS Indicated Pwr, W	IMS Indicated Pwr, dBm	dB diff, Spec An & IMS
0	5.43	3.49	3	34.77	-0.66
1	8.44	6.98	6	37.78	-0.66
2	10.3	10.72	10	40.00	-0.30
3	11.51	14.16	14	41.46	-0.05
4	14.22	26.42	27	44.31	0.09

Table 11 - IMS Forward Power Calibration

Reflected Power					
Transmitter Pwr Setting (0 to 10 scale)	Spec An Pwr, dBm	Spec An Pwr, W	IMS Indicated Pwr, W	IMS Indicated Pwr, dBm	dB diff, Spec An & IMS
0	5.4	3.47	3	34.77	-0.63
1	8.72	7.45	7	38.45	-0.27
2	10.61	11.51	11	40.41	-0.20
3	11.91	15.52	15	41.76	-0.15
4	13.87	24.38	25	43.98	0.11

Table 12 - IMS Reflected Power Calibration

Table 13 compares SWR values measured using the Anritsu vector network analyzer and the IMS SWR meter.

SWR Measurement: IMS versus Anritsu Network Analyzer	
Anritsu MS4622B VNA	IMS SWR Meter
1.08	1.06
2.00	1.94
3.02	3.08
4.01	4.12
5.01	5.25
24.7	24

Table 13 - IMS SWR Meter Comparison to VNA

6. Matching Network

To test automatic matching ability, a test of five trials was conducted for 8 different frequencies using the following settings:

The test was performed using a 50 Ω resistive dummy load and 4:1 balun at the impedance matching system antenna connector. Figure 43 shows S11 for the dummy load and balun combination from 10 MHz to 30 MHz.

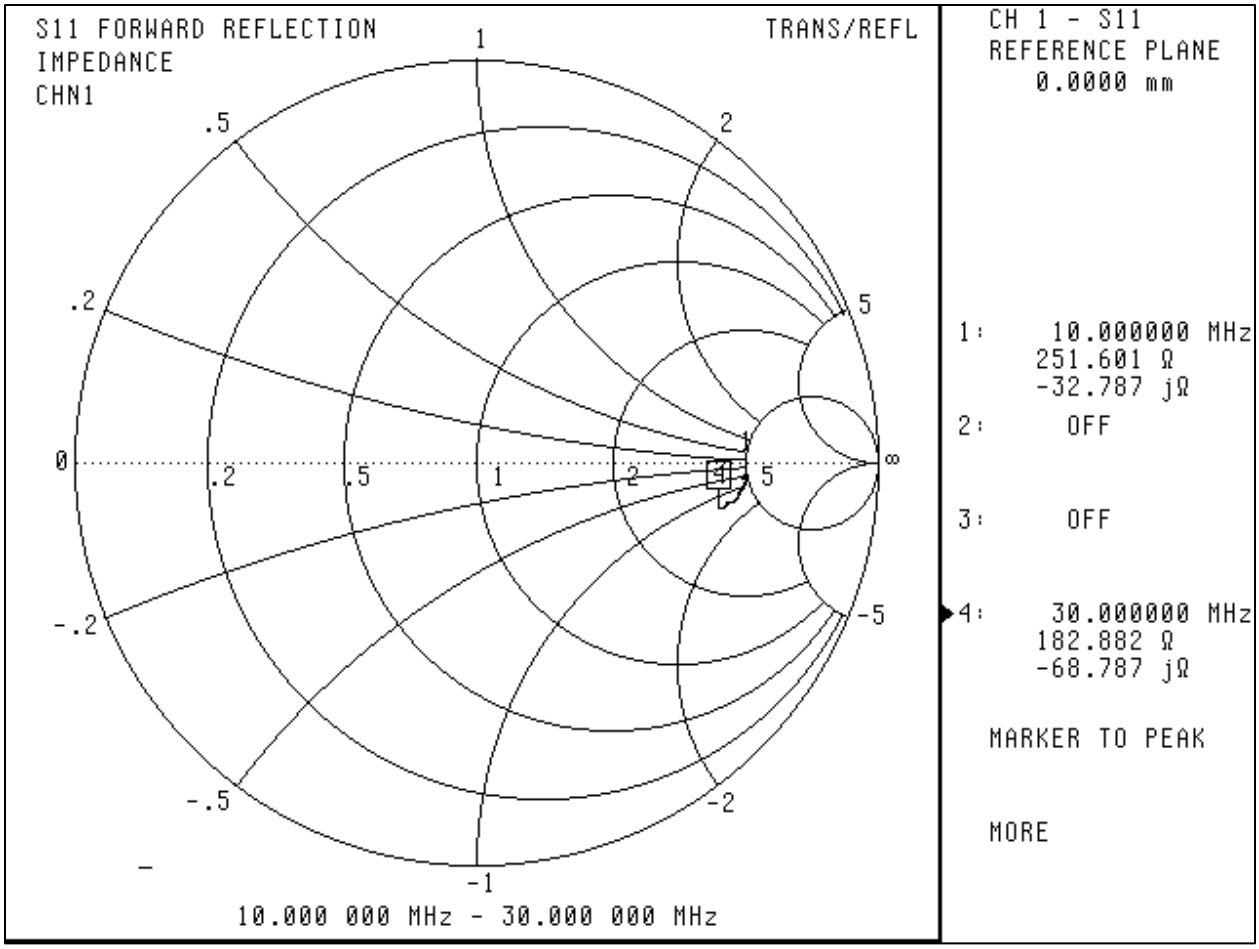

Figure 43 - S11 for 50 Ω load and 4:1 Balun.

Table 14 shows results for all test trials. The transmitter power was set at 10 W. SWR was measured using the IMS internal SWR meter.

	SWR results for IMS Dummy Load Matching Ability				
Frequency, MHz	Trial 1	Trial 2	Trial 3	Trial 4	Trial 5
3.500	1.46	1.35	1.40	1.40	1.46
7.100	1.04	1.04	1.43	1.29	1.15
10.125	1.43	1.35	1.19	1.29	1.19
14.250	1.24	1.29	1.35	1.32	1.40
18.127	1.27	1.35	1.22	1.15	1.24
21.200	1.43	1.22	1.10	1.22	1.29
24.930	1.43	1.29	1.43	1.15	1.17
28.850	1.19	1.19	1.24	1.32	1.24

Table 14 - Results of IMS match testing into dummy load

As shown in Table 14, the system provided SWR values of less than 1.5:1 for 100% of the trials at all frequencies.

The IMS was then tested for dipole antenna matching ability. The dipole used for testing utilizes the Cal Poly Amateur Radio Club antenna tower for the dipole center support approximately

40 ft above ground. The dipole ends are tied off using nylon rope to points on the EE building roof and are approximately 25 ft above the ground. Figure 44 shows the dipole location.

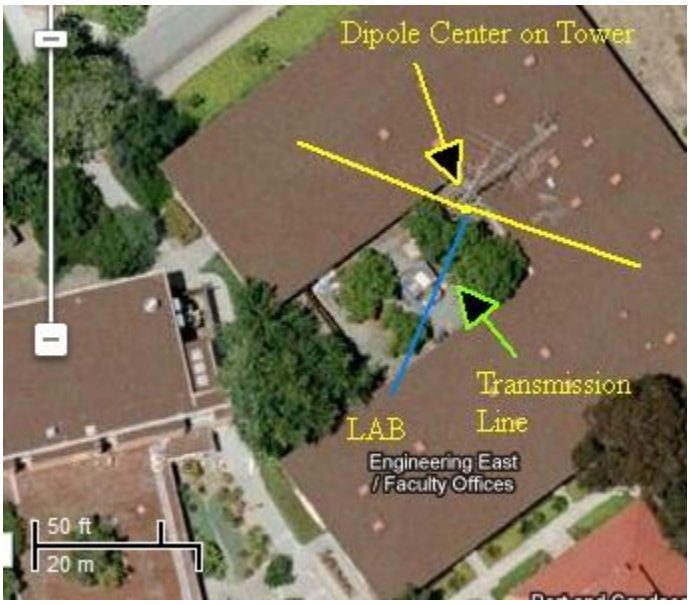

Figure 44 - Dipole Antenna Location

The dipole is connected to the IMS using 90 ft of 450 Ω balanced transmission line. Balanced transmission line is available with 300 Ω, 450 Ω, and 600 Ω characteristic impedance and 450 Ω line was an arbitrary choice.

The dipole length was originally 140 ft but was reduced to 120 ft in an attempt to reduce resistance and reactance in the 14 MHz Amateur radio band. This is the most heavily used band during daylight hours. The IMS requirement was ability to match a dipole for all HF frequencies above its half-wave resonant frequency. Under this requirement, antenna length adjustment should not be necessary. Testing determined that this requirement would not be attainable with the system as designed. Operation on 14 MHz was desired so an attempt was made to adjust the antenna length to allow matching in this band. Figure 45 compares the dipole impedance for 120 ft and 140 ft lengths. The length adjustment lowered the 14 MHz resistance and reactance.

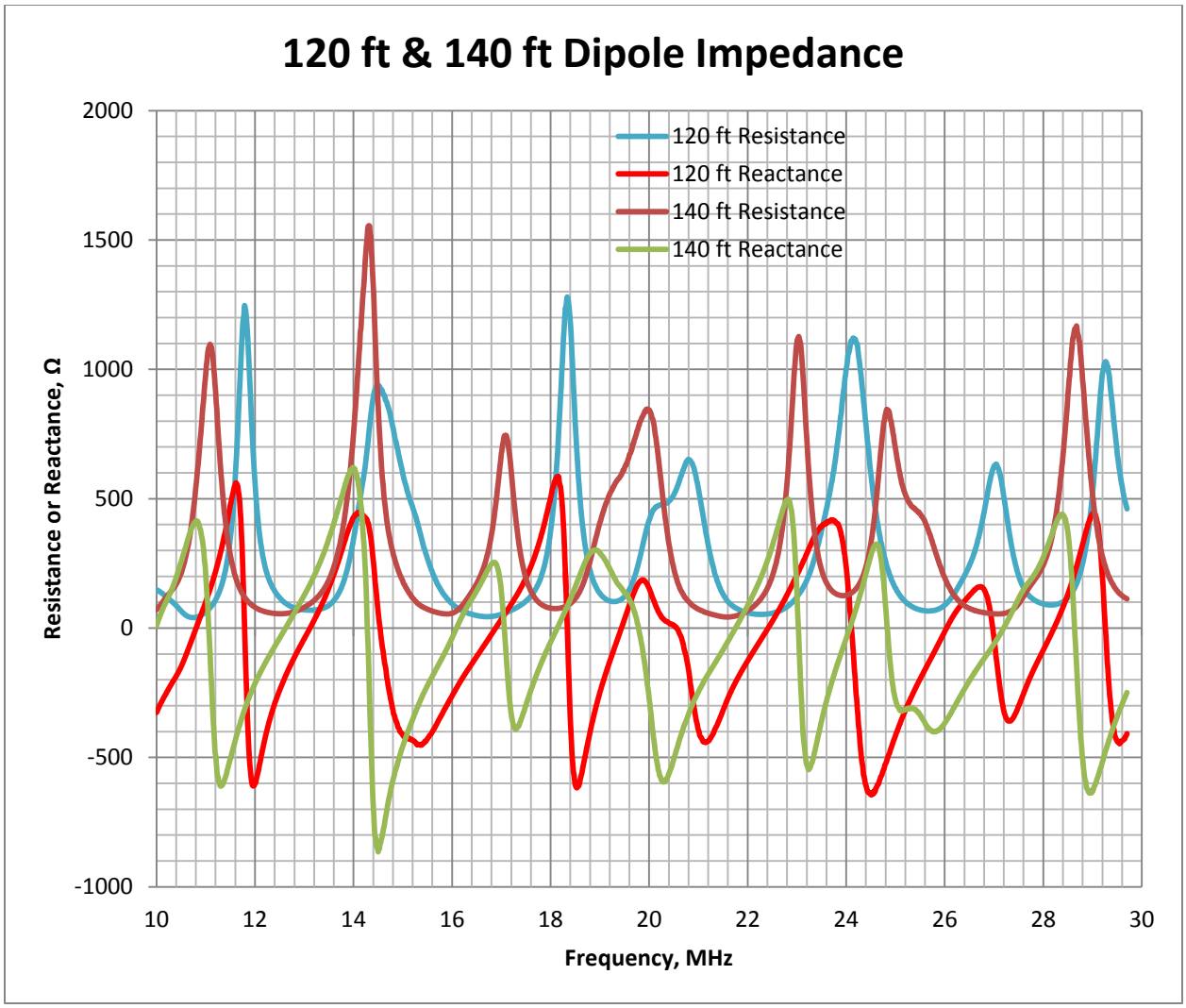

Figure 45 – 120 ft & 140 ft Dipole Impedance

Figure 46 shows the Smith chart plot of the 120 ft dipole impedance measured with an HP8753A VNA. The figure shows impedance mostly inside the r = 1 circle. Impedance outside the r = 1 circle is mostly capacitive allowing use of the Series C – Shunt L matching circuit. From 7.3 to 8 MHz, the impedance is outside r = 1 and inductive, requiring the Shunt L – Series C circuit.

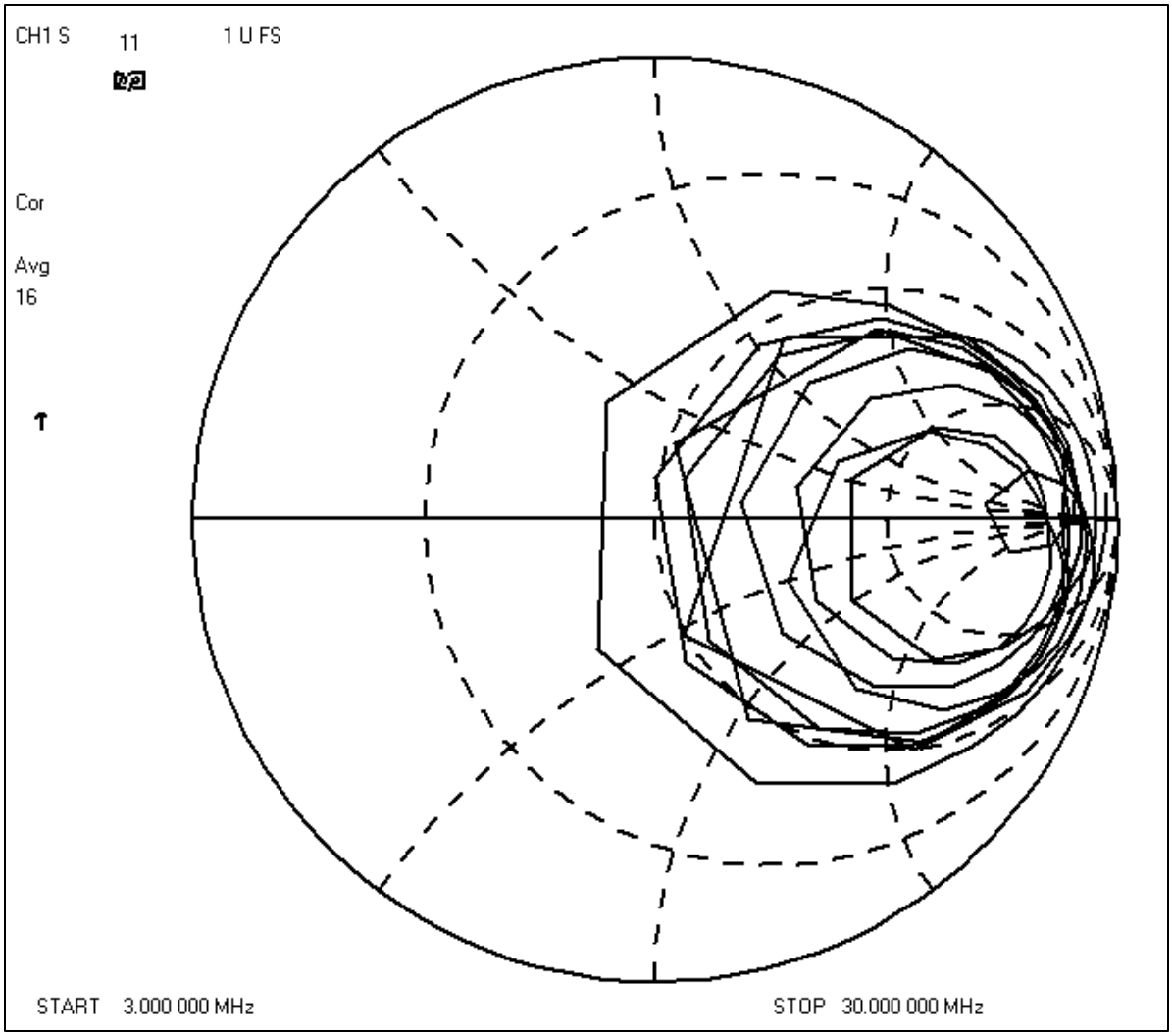

Figure 46 – 120 ft Dipole S11

Figure 47 shows the 120 ft dipole impedance from 3 – 30 MHz.

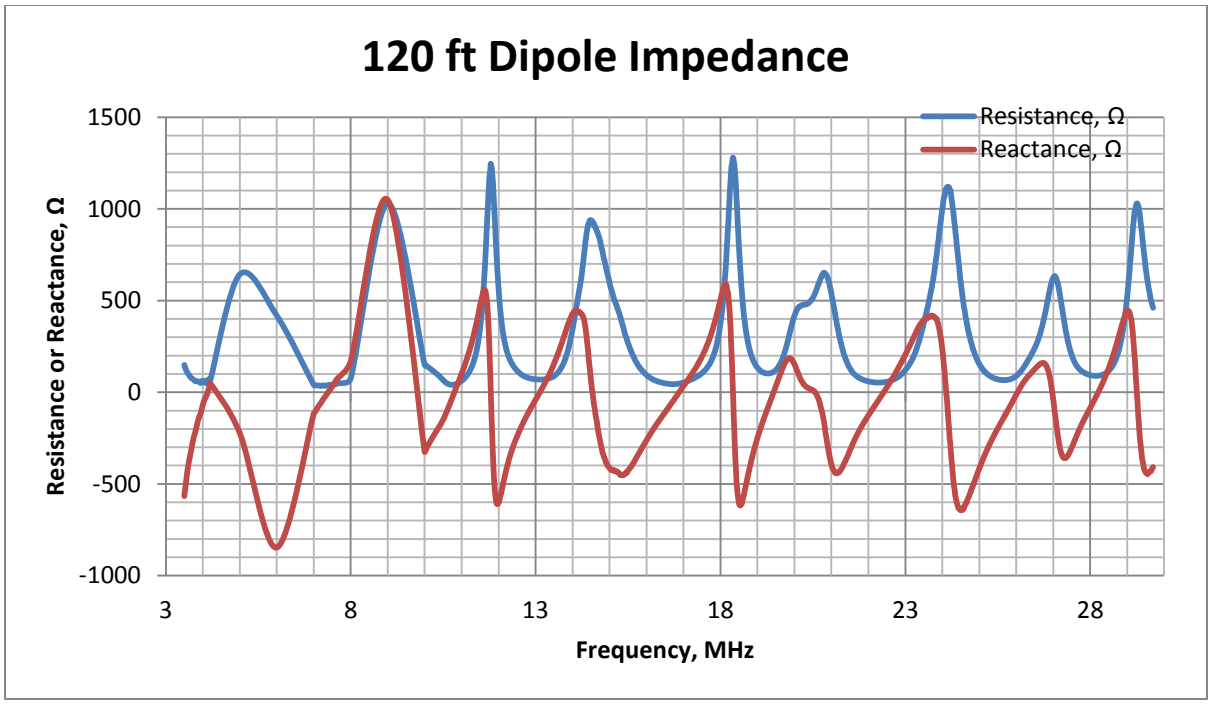

Figure 47 – 120 ft Dipole Impedance, 3 – 30 MHz

To find the IMS matching capability range, the operating frequency was set to several frequencies within all Amateur Radio bands from 3 – 30 MHz, shown in Table 6.

Automatic tuning was attempted on each frequency. If the IMS achieved a match of <1.5 SWR, the dipole impedance for that frequency was marked with an X on the Smith Chart of Figure 47. A green border encloses the matching capability impedance range.

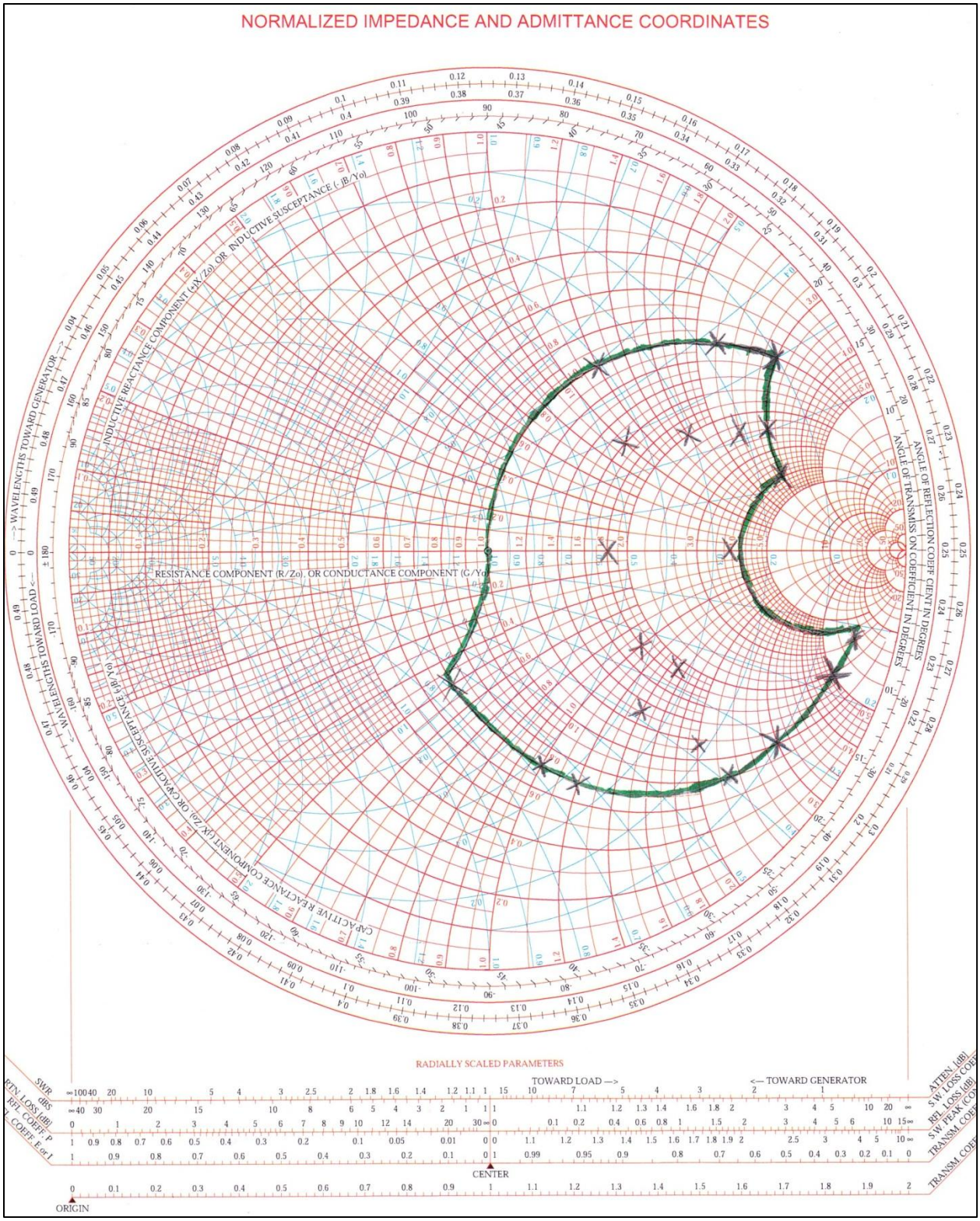

Figure 48 - Smith Chart Plot of IMS Matching Capability

Commercial automatic tuner manufacturers give tuning range specifications either in maximum SWR matching capability or impedance magnitude,| Z|, range. Equation (23) yields impedance magnitude from complex impedance, R+jX.

$$|Z| = \sqrt{R^2 + (jX)^2} \qquad (23)$$

Commercial automatic tuners have SWR matching ranges from 3:1 to 32:1. Impedance magnitude matching range is 4 – 1600 Ω [1] [2] [3] [25]. The IMS matching range is 26:1 SWR and 40-520 Ω impedance magnitude.

Since the IMS is not capable of measuring load impedance, it cannot determine the required L network configuration. The matching network switch was designed with the assumption that the IMS can match any possible load. Since that is not possible, automatic switch actuation is not enabled. If enabled, the system could fail at a matching attempt for a load impedance inside the r = 1 circle and attempt to find a match using the other network configuration, which would also fail. The tuning cycle takes approximately 100 seconds to reach the failed attempt. With automatic switching enabled, the cycle becomes 200 seconds.

Chapter 8. Conclusion

The IMS met all but one requirement specified in Table 2. It cannot match a 140 ft dipole for all frequencies between 3.5 to 30 MHz. The antenna impedance for a given frequency must fall within the range shown in Figure 48. The minimum SWR matching technique does not work for all possible load impedances since this method does not move the impedance to the r = 1 circle with subsequent capacitance adjustment to obtain the match. In Figure 48, for cases where the capacitor reactance is set ~ 0, adjusting inductance for minimum SWR places the source side impedance on x = 0 line. With x = 0, adjusting capacitance moves the impedance along constant r circle, keeping resistance constant and changing reactance. For this case, the impedance can never reach $\Gamma = 0$. The inductor must be set so the impedance approaches the r = 1 circle. The L network as constructed is capable of providing matching on all frequencies if manually tuned by trial and error. Suggested improvements include operating frequency and complex load impedance measurement. With these abilities, the tuner could attain an impedance match with only two motor adjustments. It would adjust the required element to reach the r = 1 or g = 1 circle by measuring the resistance and stopping where it equals 50 Ω, then adjust the other element to reach the match. At 40 to 100 seconds, the IMS tuning method is slow compared to commercial tuners which work in less than 15 seconds [25]. This requires the lowest possible transmitter power of 3 W to minimize amplifier heating from reflected power. The IMS does allow dipole use over a wider frequency range than otherwise possible making it a useful device for HF radio operation. Table 15 compares IMS specifications to commercial tuners.

IMS vs. Commercial Antenna Tuner Specifications		
	Commercial Tuners [1] [2] [3] [25]	IMS
Maximum SWR	3:1 to 100:1	26:1
Impedance Magnitude Range	4 – 1600 Ω	40 – 520 Ω
Tuning Time	0.5 – 15 seconds	40 – 100 seconds

Table 15 - IMS vs. Commercial Antenna Tuner Specifications

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Chapter 10. Appendices

Appendix A: Senior Project Analysis

Project Title:
1.8 to 30 MHz Automatic Balanced Transmission Line Antenna Impedance Matching System

Student's Name:
William Blodgett

Student's Signature:

Advisor's Name:
Professor Dean Arakaki

• 1. Summary of Functional Requirements

The Impedance Matching System (IMS) connects in line with the transmission line between a radio transmitter and the antenna. It allows antenna use over a wider bandwidth than otherwise possible. It senses reflected power due to transmitter-antenna impedance mismatch and varies its internal impedance to achieve a match, thereby minimizing reflected power.

• 2. Primary Constraints

One challenge is minimizing project cost. There are not many part sources due to the parts' specialized nature such as high voltage variable capacitors. Due to lower economy of scale, these parts are very expensive.
A second challenge is the overall project complexity and required knowledge in subject areas beyond the Cal Poly EE curriculum. Subject areas include transmission line power measurement and stepper motor control systems. Fortunately, the curriculum provided enough basic knowledge to begin design and development.

• 3. Economic

The IMS required approximately 300 man-hours to design and build. The project also resulted in increased activity of more people than simply the person designing and building it. The parts require manufacturing, sales, and delivery. Others are involved in the production of raw

materials required to make the parts such as metals, silicon, and plastic, which are also natural capital. The IMS benefits people who use it by facilitating their radio communications efforts.

The project involved money exchange between the project owner, parts distributers, and manufacturers. If the project will be marketed to consumers, money exchange will be involved there as well.

The project required various manufactured capital items such as tools, electronic test equipment, vehicles for transportation, and computers. A vector network analyzer is the most useful tool for testing the IMS. A VNA costs $10,000 or more. Fortunately, the university makes them available for student use. A 100 W HF transceiver, typically around $1000, was used for operational testing. An accurate inline power meter ($150) was required for IMS internal power meter calibration.

All development costs accrued between September 2011 and May 2012. Manufacturing costs will accrue after this timeframe. The project owner pays these costs.

See Appendix B for a Gantt chart detailing the project schedule and Appendix C for a detailed listing of parts and development costs.

Since the IMS is an electronic device with moving parts, it will not continue to work indefinitely. Based on personal experience, the IMS is estimated to operate from 2 to 20 years before failure. This depends on many variables, the most important; how often moving parts operate. The IMS requires some periodic user maintenance, such as lubricating moving parts. Even after the IMS is no longer sold to consumers, the manufacturer should maintain support and replacement parts for the product.

• 4. If manufactured on a commercial basis:

An order of magnitude estimate for unit sales per year is 1000 based on 2 million licensed amateur radio operators worldwide. However, a small percentage would likely find the IMS affordable and would choose to buy it.

As detailed in Appendix C, the cost estimate for a prototype is $716 for parts and required approximately 300 hours to develop. At $25/hour labor, the total development cost is $8216. With the working prototype developed, the time to assemble one unit could decrease to five hours. Assuming a fixed parts cost of $716, the estimated total manufacturing cost for one unit is $841 based on $25 labor cost.

Similar antenna tuner products sell for $1500. If the IMS sells for this price, the profit is $659 per unit. The estimated gross profit per year is $659 x 1000 = $659,000. Subtracting development cost results in $650,784. This does not include all other business costs such as

taxes, insurance, employing additional workers, property, and other expenses. A true profit estimate requires a more detailed analysis far beyond the scope of this report.

The primary operating cost for the IMS is electricity. A rough estimate of the average IMS power consumption is 15 W. The California Public Utilities Commission website (http://www.cpuc.ca.gov) shows the current average electricity rate at $0.16/kWh. The electricity cost for operating the IMS is approximately 16 cents for 67 hours of operation.

• 5. Environmental

Manufacturing the IMS requires natural resources such as metals, silicon, and petroleum products such as plastic. Gathering these resources requires burning fuel, which contributes to air pollution. Mining and drilling for oil can also cause water pollution and disturb natural habitats for many species. An example is the 2010 BP Gulf of Mexico Oil Spill. Manufacturing and use of the IMS also requires electricity. Generating electricity creates air pollution and requires mining and drilling for fuel sources.

As detailed in the book *Computers and the Environment: Understanding and Managing Their Impacts*, a microchip, for example, requires 2.9 kWh of electricity and 970 grams of fossil fuel to produce. [26]

• 6. Manufacturability

Material expenses for IMS construction could be less if more of the parts, such as variable capacitors, were built in-house instead of ordered from a third party. However, building a variable capacitor from metal stock could require a significant initial investment in tools such as a stamp press to cut metal plates. If all parts including printed circuit boards are sourced from third parties, manufacturing would only involve simple soldering and machining using consumer level tools.

• 7. Sustainability

Initially, recyclable parts and materials may not be selected. This contributes to unsustainable consumption of raw material resources. Manufacturing could make use of innovative techniques to support sustainability. The challenge is that there are not many practices fully in place to make electronic devices fully recyclable. E-waste programs are ubiquitous to keep electronics out of landfills, but full recycling may require several years. According to the United Nations initiative, "Solving the E-waste Problem" (http://www.step-initiative.org/):

“Because of this complex composition of valuable and hazardous substances, specialized, often “high-tech” methods are required to process e-waste in ways that maximize resource recovery and minimize potential harm to humans or the environment. Unfortunately, the use of these specialized methods is rare…” [27]

Based on this, the IMS may not be 100% recyclable and sustainable until the electronics industry as a whole makes further progress toward full sustainability.

• 8. Ethical

The IEEE Code of Ethics [28], the Golden Rule, and Kant’s Categorical Imperative provide guidance in considering ethical implications involving the IMS project.

One possible unethical use of the IMS involves unlicensed or otherwise unlawful radio transmissions. There have been many cases of individuals deliberately engaging in malicious activities involving radio equipment. For example, people have deliberately used radio equipment to cause harmful interference or “jam” law abiding radio users. Unfortunately, there are no laws in place to stop such individuals from gaining possession of radio equipment. Fortunately, these individuals are often caught using radio direction finding equipment.

Malicious radio jamming clearly violates the Golden Rule and, more universally speaking, the Categorical Imperative. If all radio operators jammed everyone else on the radio, then radio would not be a useful communications tool. Jamming also violates Rules 1 and 9 of the IEEE Code of Ethics. It is injurious to others and can endanger “safety, health, and welfare of the public” by interfering with public safety communications. [28]

An ethical implication involving IMS design is intellectual property use. Since the IMS design incorporates unoriginal circuitry, in actuality, it will never be sold for profit since it would be unethical to profit from others’ intellectual property without their consent. The Golden Rule applies since most people would not choose to allow others to profit from their ideas. Rule 7 of the IEEE Code of Ethics says “to credit properly the contributions of others.” [28]

A positive ethical IMS aspect is its utility for emergency and international communications. As described in the Social and Political Section, it could facilitate disaster communications and communications across borders, which support IEEE Code of Ethics Rule 1 and the Golden Rule.

Last, but not least, in accordance with IEEE Code of Ethics Rules 1 and 9 and the Golden Rule, the chief consideration in the IMS design will be safety. Hazards such as fire and electric shock will be mitigated to the fullest extent practical.

• 9. Health and Safety

IMS manufacture could potentially expose workers to hazardous materials such as solder fumes, particulate matter from machining, and chemical solvents used for removing solder flux. Proper ventilation and appropriate personal protective equipment such as safety goggles and dust masks help mitigate these hazards.

Using the IMS poses some electrical hazards. It requires 120 V AC household power, which is lethal. The closed chassis protects users from this hazard provided that they do not open the chassis while the IMS is energized. The radio frequency energy from a 100 W transmitter also poses a hazard. RF energy is known to cause burns. The exposed balanced transmission line terminals on the exterior of the chassis carry RF energy at high voltage. The user and others must avoid touching them during transmitter operation.

• 10. Social and Political

The IMS could greatly benefit disaster affected communities. Major disasters often damage, overload, or otherwise limit the public telephone network availability. This hinders relief efforts. Two-way radio often provides the most reliable communications system in such situations. If the disaster is widespread, the only way to relay communications into and out of a region may be via satellite or HF radio. The IMS can facilitate disaster communications by allowing a user to set up one simple wire antenna and operate on multiple frequency bands.

In everyday Amateur Radio use, it can facilitate international communication between operators. Open dialog between people in different countries benefits international relations because people are able to understand and relate to each other better when they talk to each other rather than blindly speculate about each other.

Inequities are created because few people can afford such a device and its associated radio equipment. Also, raw materials are often produced in third world countries where workers may not profit fairly compared to the companies taking advantage of their low-cost labor.

• 11. Development

As mentioned in the Primary Constraints section, this project required advanced electrical theory beyond that taught in the Cal Poly EE curriculum. For example, the project required research into techniques and circuitry used to measure power on a transmission line. A common circuit for power measurement is a directional coupler, which provides a proportionally reduced transmission line energy sample measured using sensitive low power devices such as diodes. An additional circuit produces a DC voltage linearly proportional to forward and reflected power detected using a non-linear diode described in the ARRL Handbook [7]. Kaune [11] provides a simplified technique using an application specific integrated circuit (ASIC). Analog Devices produces a logarithmic amplifier, the AD8307, which produces a DC output voltage linearly proportional to an input radio frequency signal in dBm up to 500 MHz. These are circuits which are not included in Cal Poly EE coursework, will be very useful to the project.

Appendix B: Time Schedule Allocation

Figure 49 shows the overall IMS project schedule presented in Gantt chart form.

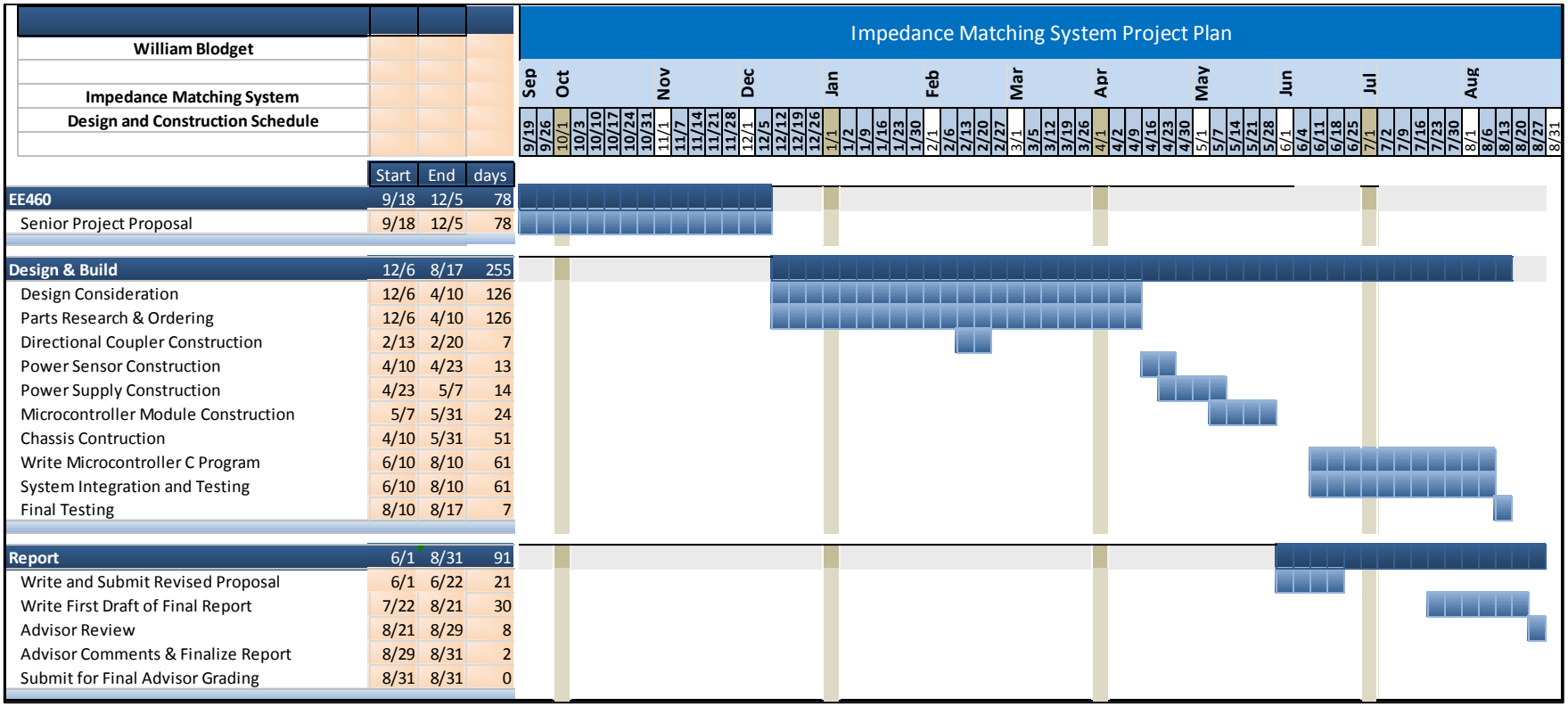


Figure 49 - IMS Project Gantt Chart

Appendix C: Parts List and Costs

Impedance Matching System Bill of Materials							
Part name	**Purpose**	**Manufacurer**	**Source**	**Part Number**	**Quantity**	**Each**	**Extended**
19" Rack metal enclosure	chassis		Salvaged		1	$ -	$ -
1/8" Aluminum Sheet Metal , Cut to size	Front, rear, bottom Panels		McCarthy Steel		1	$ 54.51	$ 54.51
ROLLER INDUCTER, 33uH	matching network	MFJ	MFJ	404-1052	1	$ 59.00	$ 59.00
700 pF Variable Capacitor	matching network		Salvaged		1	$ 20.00	$ 20.00
Insulator, ceramic feedthrough	antenna terminals	DX Engineering	DX Engineering	DXE-9551	2	$ 11.95	$ 23.90
ATmega32 microcontroller, 40 PDIP	System Control	Atmel	Digikey	ATMEGA32-16PU	1	$ 8.80	$ 8.80
40 pin DIP IC Socket	Microcontroller mounting	Sparkfun	Sparkfun	PRT-07944	1	$ 0.95	$ 0.95
PC Board, 2-3/4x3-1/16	Microcontroller mounting	Radio Shack	Radio Shack	276-158	1	$ 3.19	$ 3.19
Crystal Oscillator kit (used 16 MHz, Radial Lead)	Microcontroller Clock	Sparkfun	Sparkfun	COM-09994	1	$ 4.95	$ 4.95
resistor, axial lead, 10kΩ, 1/4W, 5%	MCU reset Pull Up		Digikey	CF14JT10K0	1	$ 0.08	$ 0.08
resistor, axial lead, 330Ω, 1/4W, 5%	MCU reset current limit		Digikey	CF14JT330R	1	$ 0.08	$ 0.08
capacitor, ceramic, 22 pF, radial lead	Microcontroller Clock	TDK	Digikey	FK18C0G1H220J	2	$ 0.20	$ 0.40
capacitor, ceramic, 100 nF, radial lead	MCU & Power Supply	TDK	Digikey	FK11C0G1H104J	8	$ 0.76	$ 6.08
ferrite beads, for #20 AWG wire, 12 pack	Microcontroller RFI Suppr.	Philmore	Fry's Electronics	86-157	1	$ 0.99	$ 0.99
header pins, breakaway, male, 2x40, 0.1"	Control Unit	Molex	Mouser	10-89-7801	1	$ 6.12	$ 6.12
header pins, breakaway, male, 1x40, 0.1"	Control Unit		Sparkfun	PRT-00116	1	$ 1.50	$ 1.50
Stepper Motor Driver	matching network control	Sparkfun	Sparkfun	ROB-10735	3	$ 22.95	$ 68.85
Heatsink, for Stepper Motor Driver IC, 5 pack	matching network control	Sparkfun	Sparkfun	PRT-10746	1	$ 7.95	$ 7.95
Stepper Motor, 2 Amp	matching network control	Sparkfun	Sparkfun	ROB-10847	2	$ 24.95	$ 49.90
Stepper Motor, 700 mA, linear actuator	matching network control	Sparkfun	Sparkfun	ROB-10848	1	$ 29.95	$ 29.95
16x2 Character LCD Display	System Status Display	Sparkfun	Sparkfun	LCD-00255	1	$ 13.95	$ 13.95
potentiometer, 10kΩ	LCD Contrast Control	NTE	Fry's Electronics	500E-0201	1	$ 2.99	$ 2.99
diode, silicon, rectifier, 1N4001	LCD Backlight Volt. Regul.	Radio Shack	Radio Shack	276-1101	1	$ 1.29	$ 1.29
DB-15 Male Connector	control unit		Radio Shack		1	$ 2.29	$ 2.29
DB-15 Female Connector	control unit		Radio Shack		1	$ 2.29	$ 2.29
DB-25 Male Connector	control unit		Radio Shack		1	$ 2.29	$ 2.29
DB-25 Female Connector	control unit		Radio Shack		1	$ 2.29	$ 2.29
DB-15 Connector Metal Hood	control unit		Radio Shack	276-1508	1	$ 3.19	$ 3.19
Power Transformer, 2 Amp, 120V/25.2CT	Power Supply		Radio Shack	273-1512B	1	$ 12.99	$ 12.99
Voltage Regulator, 5 V, 1.5 A Fixed, TL780-05	Power Supply	TI	Mouser		1	$ 0.77	$ 0.77
Voltage Regulator, 12 V, 2 A Fixed, L78S12	Power Supply	ST	Mouser		1	$ 0.78	$ 0.78
Capacitor, 4700 uF, 50 V	Power Supply	Nichicon	Mouser	UVZ1H472MRD	2	$ 1.78	$ 3.56
Rectifier, Bridge, 4 A, 400V PIV, TL404G	Power Supply		HSC		1	$ 1.90	$ 1.90
Fuse, 1 Amp, glass, 5 pack	Power Supply	Buss	Home Depot		1	$ 1.85	$ 1.85
plug, RCA phono, male, 4 pack	Power Supply	Radio Shack	Radio Shack	274-0319	1	$ 3.19	$ 3.19
PC Board, 1-27/32x2-27/32"	Power Supply	Radio Shack	Radio Shack	276-149	1	$ 2.19	$ 2.19
capacitor, ceramic, 1 nF, radial lead	Power Supply	TDK	Digikey	FK14C0G2E102J	1	$ 0.29	$ 0.29
capacitor, ceramic, 220 nF, radial lead	Power Supply	TDK	Digikey	FK28X5R1E224K	1	$ 0.34	$ 0.34
Switch, pusbutton, normally open, 4 pack	control	Radio Shack	Radio Shack	275-1547	1	$ 3.69	$ 3.69
AD8307 surface mount log amp	RF Power Sensor	Analog Devices	Digikey	AD8307ARZ-ND	2	$ 13.16	$ 26.32
100 uH RF Choke Inductor	RF Power Sensor		Mouser		2	$ 0.18	$ 0.36
Resistor, SMT, 52.3Ω, 1W, 1%, 2512 Size	RF Power Sensor		Digikey	PT52.3AFCT-ND	2	$ 0.94	$ 1.88
PCB, RF Power Sensor	RF Power Sensor	Far Circuits	Far Circuits		1	$ 25.00	$ 25.00
Capacitors, SMT, kit for RF Sensor	RF Power Sensor	Far Circuits	Far Circuits		1	$ 3.00	$ 3.00
BNC female PCB connector	RF Power Sensor	TE Connectivity	Digikey	1-1634612-0	2	$ 2.62	$ 5.24
AWG#14 Magnet Wire	balun	Amidon	Amidon		20	$ 0.40	$ 8.00
12 gauge Teflon tubing	balun	Amidon	Amidon		20	$ 0.70	$ 14.00
8"x10" x 0.93" Polycarbonate Sheet	balun		Home Depot		1	$ 3.98	$ 3.98
Mix 31 2.4 Inch Toroid	balun	KF7P	KF7P.com		1	$ 7.95	$ 7.95
bolt, nylon, 5/16" x 2"	balun		Ace Hardware		1	$ 0.50	$ 0.50
Nut, nylon, 5/16"	balun		Ace Hardware		1	$ 0.50	$ 0.50
Washer, nylon, 5/16"	balun		Ace Hardware		2	$ 0.25	$ 0.50
Spacer, nylon, 5/16"	balun		Ace Hardware		3	$ 0.25	$ 0.75

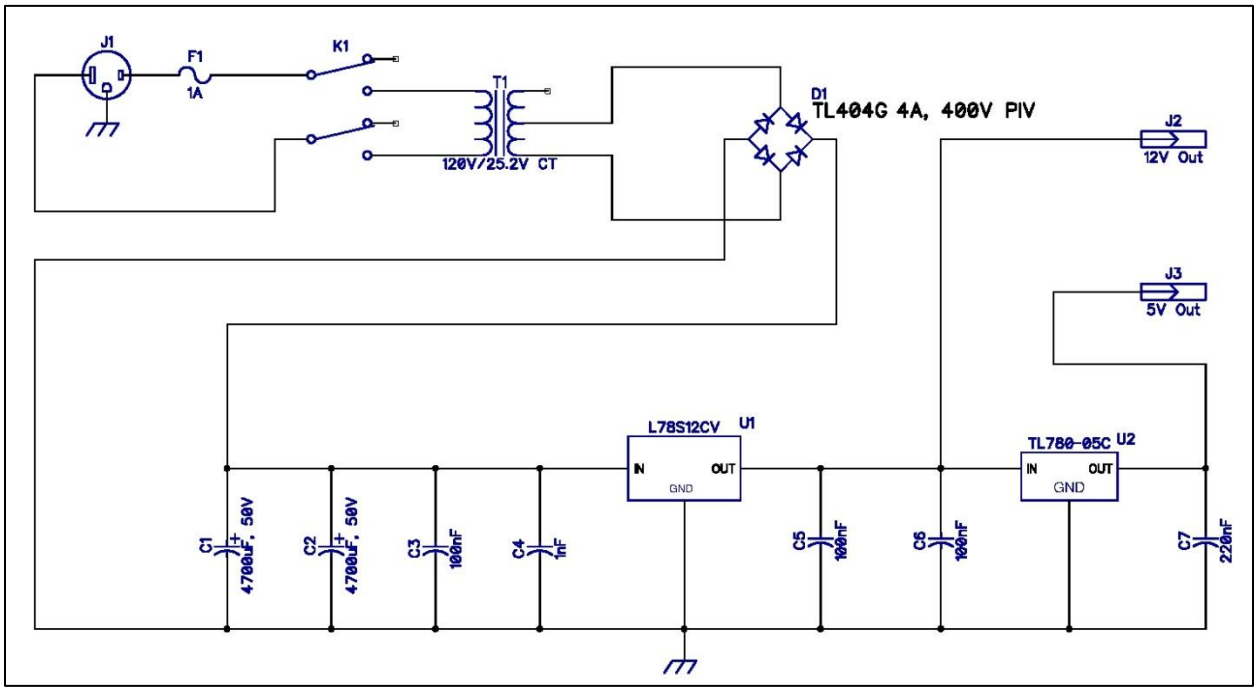
Table 16 - Antenna Impedance Matching System Bill of Materials (Page 1 of 2)

Part name	Purpose	Manufacurer	Source	Part Number	Quantity	Each	Extended
PL-259 Male UHF Coaxial Connector	RF cabling		RF Parts		3	$ 1.95	$ 5.85
Reducer, PL-259 for RG-58/U	RF cabling		RF Parts	UG175/U-S	3	$ 0.55	$ 1.65
RG-58 Coaxial Cable	RF cabling	Belden	RF Parts		20	$ 0.35	$ 7.00
BNC Male Clamp Connector	RF cabling	RF Industries	RF Parts	RFB-1100-2ST	4	$ 3.30	$ 13.20
20 dB Attenuator	Directional Coupler	Mini-Circuits	Mini-Circuits	HAT-20	2	$ 9.95	$ 19.90
RG-8 Coaxial Cable	Directional Coupler	Belden	RF Parts		1	$ 2.15	$ 2.15
SO-239 Chassis Mount Connector	Directional Coupler		Newhamstore		2	$ 1.65	$ 3.30
BNC Female Chassis Mount Conn.	Directional Coupler	RF Parts	RF Parts	RFB-1115	4	$ 3.55	$ 14.20
Double Sided Copper Clad PCB (6"x9")	Directional Coupler	Injectorall	Digikey	PC53-ND	1	$17.14	$ 17.14
Enamel Wire AWG# 26 0.25lb spool	directional coupler	Amidon	Amidon		1	$ 7.95	$ 7.95
FT-82-67 ferrite torriod	directional coupler	Amidon	Amidon		2	$ 1.20	$ 2.40
Aluminum Box	RF Sheilding	Radio Shack	Radio Shack	270-238	3	$ 3.19	$ 9.57
2-56 Machine Screws, assorted length, 42 pack	chassis	Radio Shack	Radio Shack	64-3010	2	$ 2.19	$ 4.38
2-56 Hex Nuts, 40 pack	chassis	Radio Shack	Radio Shack	64-3017	1	$ 2.19	$ 2.19
4-40 Machine Screws , assorted length, 42 pack	chassis	Radio Shack	Radio Shack	64-3011	2	$ 2.19	$ 4.38
4-40 Hex Nuts, 31 pack	chassis	Radio Shack	Radio Shack	64-3018	2	$ 2.19	$ 4.38
6-32 Machine Screws	chassis	Radio Shack	Radio Shack	64-3012	1	$ 2.19	$ 2.19
6-32 Machine Nuts	chassis	Radio Shack	Radio Shack	64-3019	1	$ 2.19	$ 2.19
Washers, flat, assorted	chassis	Radio Shack	Radio Shack	64-3022	2	$ 2.19	$ 4.38
Standoff, aluminum, 4-40 threaded, 3/8", 10 pack	control unit & pwr supply		Sparkfun	PRT-10739	1	$ 2.95	$ 2.95
#30 Wrapping Wire	control unit	Radio Shack	Radio Shack	278-502	1	$ 4.99	$ 4.99
SPST Pushbutton Switch, 4 pack	Front Panel Buttons	Radio Shack	Radio Shack	275-1547	1	$ 3.69	$ 3.69
Threaded Rod, #10-24, 3 foot long	chassis		Ace Hardware		2	$ 4.59	$ 9.18
#10-24 Nuts, 100 Pack	chassis		Ace Hardware		1	$ 7.95	$ 7.95
#10 Flat Washers, 100 Pack	chassis		Ace Hardware		1	$ 4.49	$ 4.49
Screws, Pan-Head Sheet-Metal, 35 pack	chassis	Radio Shack	Radio Shack	64-3016	1	$ 2.19	$ 2.19
Rubber Grommets, Assorted, 31-Pack	wire feedthrough	Radio Shack	Radio Shack	64-3025	1	$ 2.29	$ 2.29
Polypropylene Cable Clamps, 18-Pack	front panel cable	Radio Shack	Radio Shack	64-3028	1	$ 2.19	$ 2.19
Copper Tape, 2" x 50 ft	RF Sheilding		Sparkfun	PRT-11081	1	$14.95	$ 14.95
rod, round, aluminum, 1/8" x 12"	RF switch		Ace Hardware		1	$ 1.29	$ 1.29
tube, round, aluminum, 1/4"x0.035"x12"	RF switch		Ace Hardware		1	$ 2.49	$ 2.49
standoff, nylon, 4-40 threaded, 1" long	RF switch		Ace Hardware		1	$ 0.10	$ 0.10
Screw, 4-40 x 0.5", Nylon	RF switch		Ace Hardware		1	$ 0.10	$ 0.10
Washer, flat, 4-40, Nylon	RF switch		Ace Hardware		2	$ 0.05	$ 0.10
angle stock, aluminum, 0.5"x0.75"x36"	mounting brackets		Ace Hardware		1	$ 4.99	$ 4.99
SO-239 to SO-239 Chassis feedthrough connector	Rear Panel RF connector	Pan	Fry's Electronics	UHF-7625P	1	$ 2.29	$ 2.29
Screw-Cap Panel-Mount Fuse Holder	Rear Panel Fuse	Radio Shack	Radio Shack	270-367	1	$ 2.99	$ 2.99
Plug, male, power, IEC C14	Rear Panel AC Plug	salvaged			1	$ -	$ -
Switch, DPST	Rear Panel power switch	salvaged			1	$ -	$ -
tube, round, aluminum, 3/8"x0.049"x12"	motor shaft coupler		Ace Hardware		1	$ 3.99	$ 3.99
bumper, rubber, screw in, 4 pack	chassis feet		Ace Hardware		1	$ 2.99	$ 2.99
Screw, brass, 8-32 x 2"	Antenna & GND terminals		Ace Hardware		3	$ 0.25	$ 0.75
Nut, brass, 8-32	Antenna & GND terminals		Ace Hardware		5	$ 0.25	$ 1.25
Washer, brass, 8-32	Antenna & GND terminals		Ace Hardware		15	$ 0.10	$ 1.50
nut, wing, 8-32	Antenna & GND terminals		Ace Hardware		3	$ 0.25	$ 0.75
						total	$ 716.16

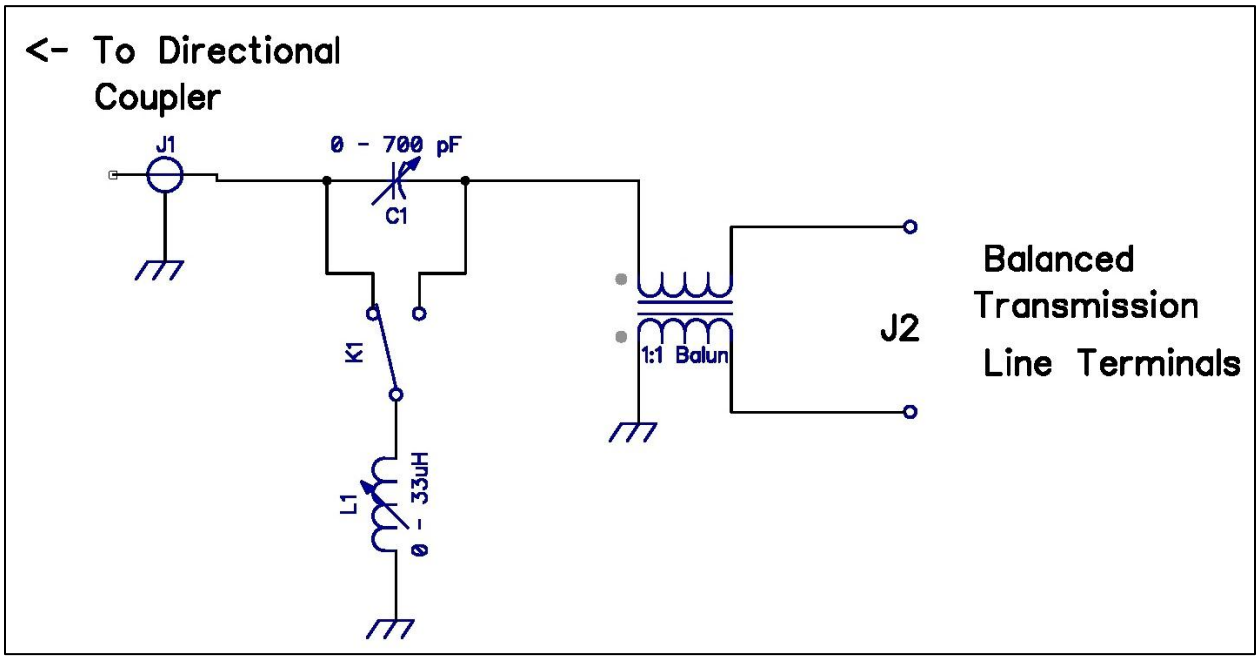
Table 17 - Antenna Impedance Matching System Bill of Materials (Page 2 of 2)

Appendix D: Schematics

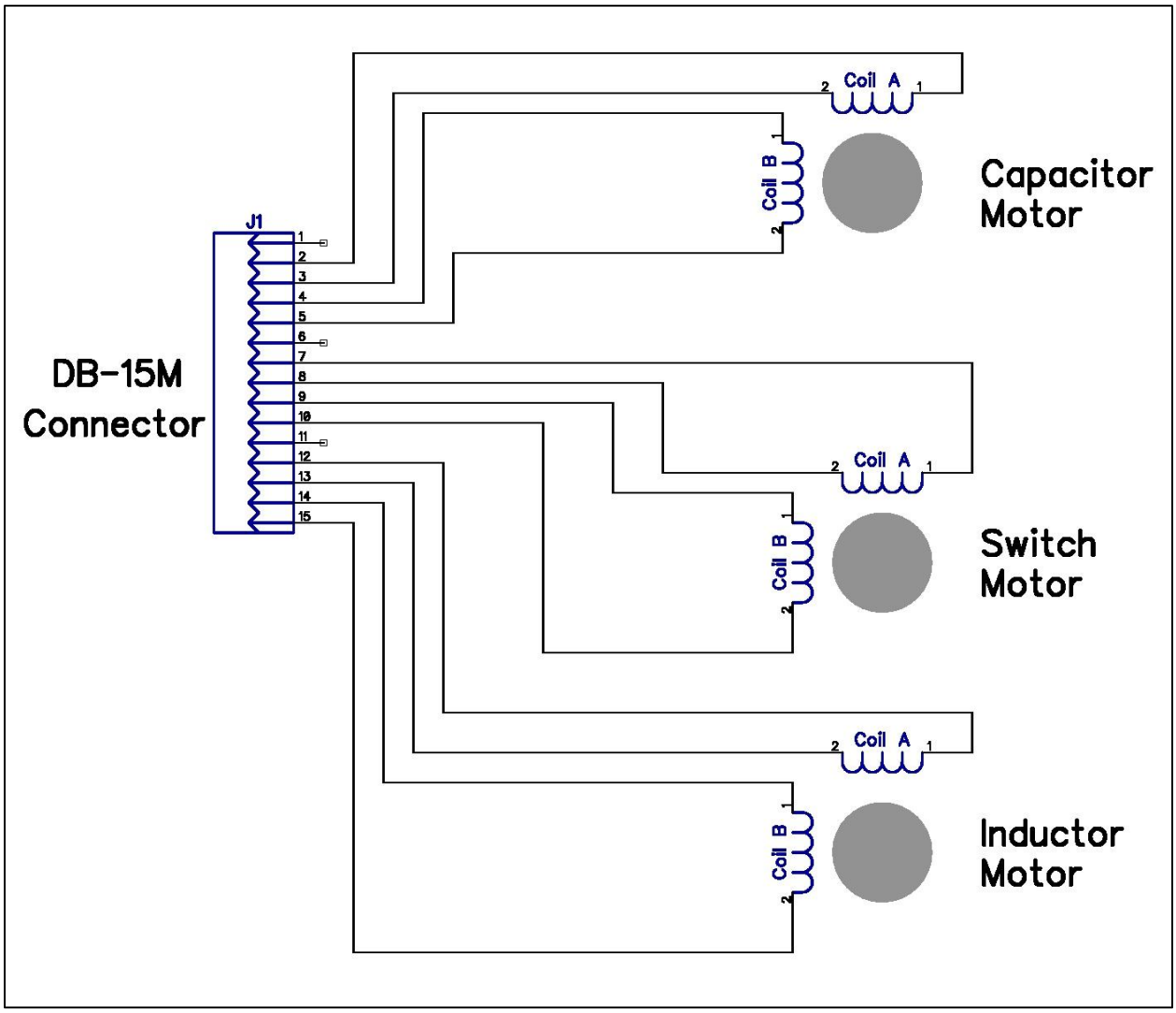
1. DC Power Supply



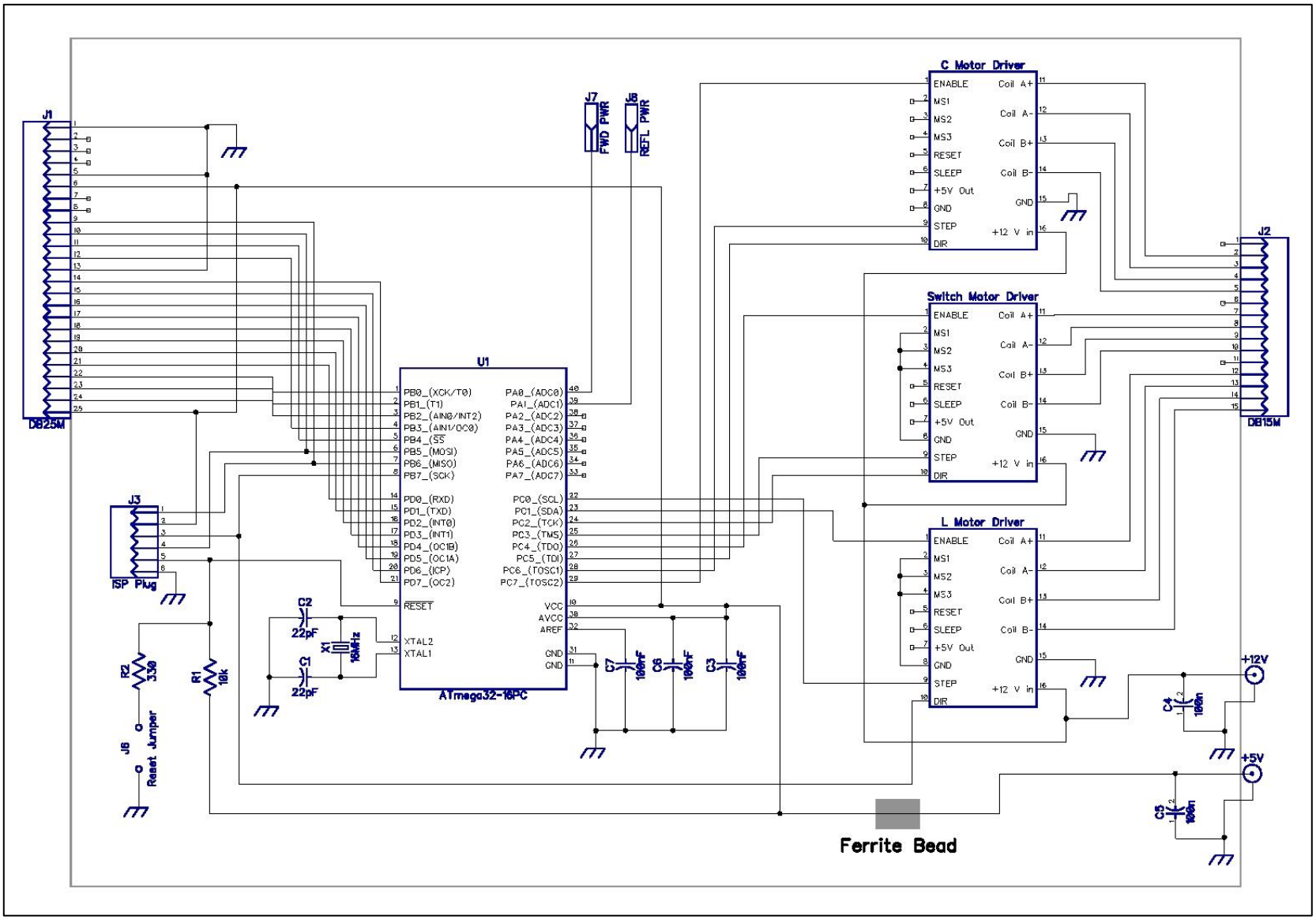
2. Matching Network



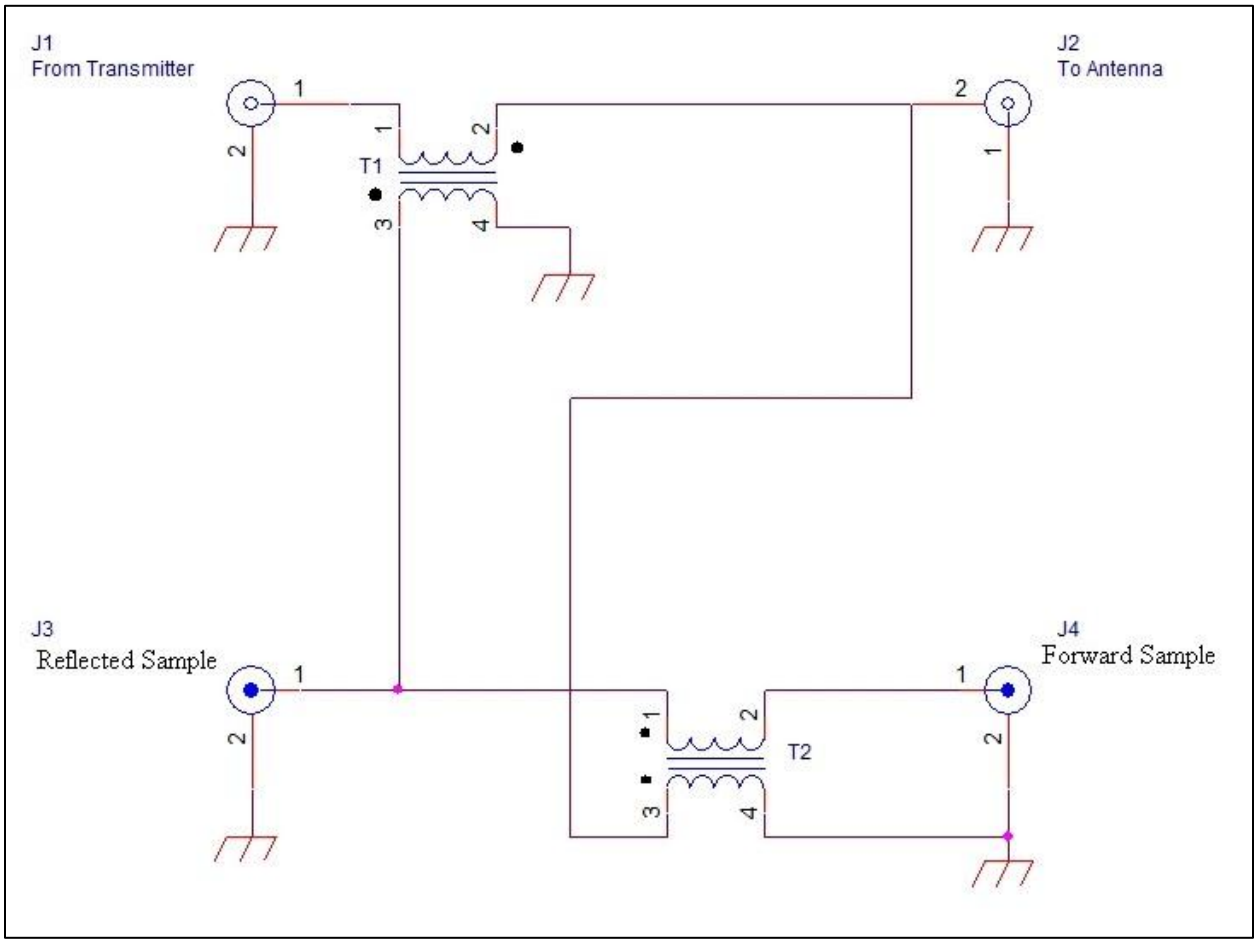
3. Stepper Motors



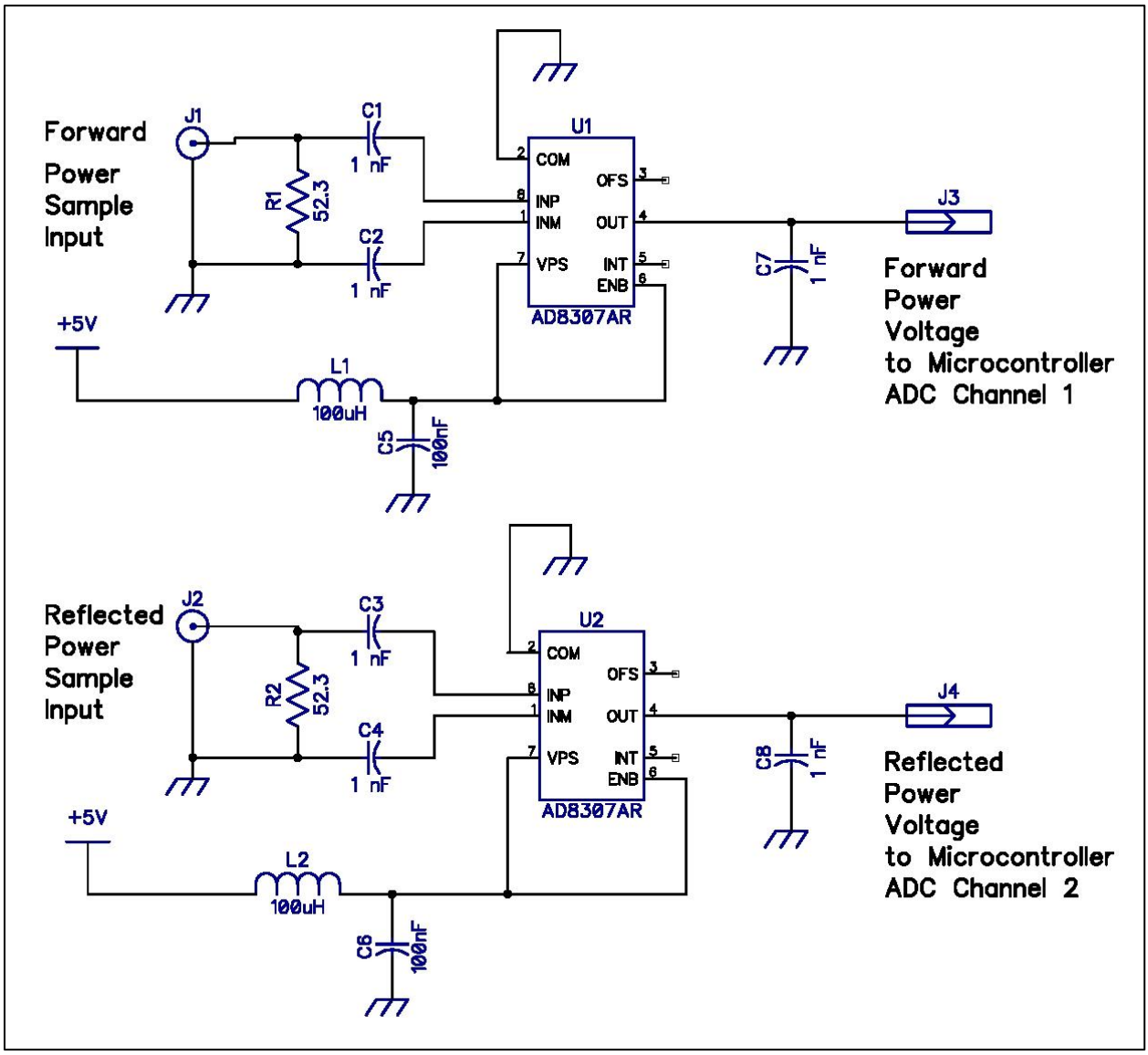
4. Control Unit



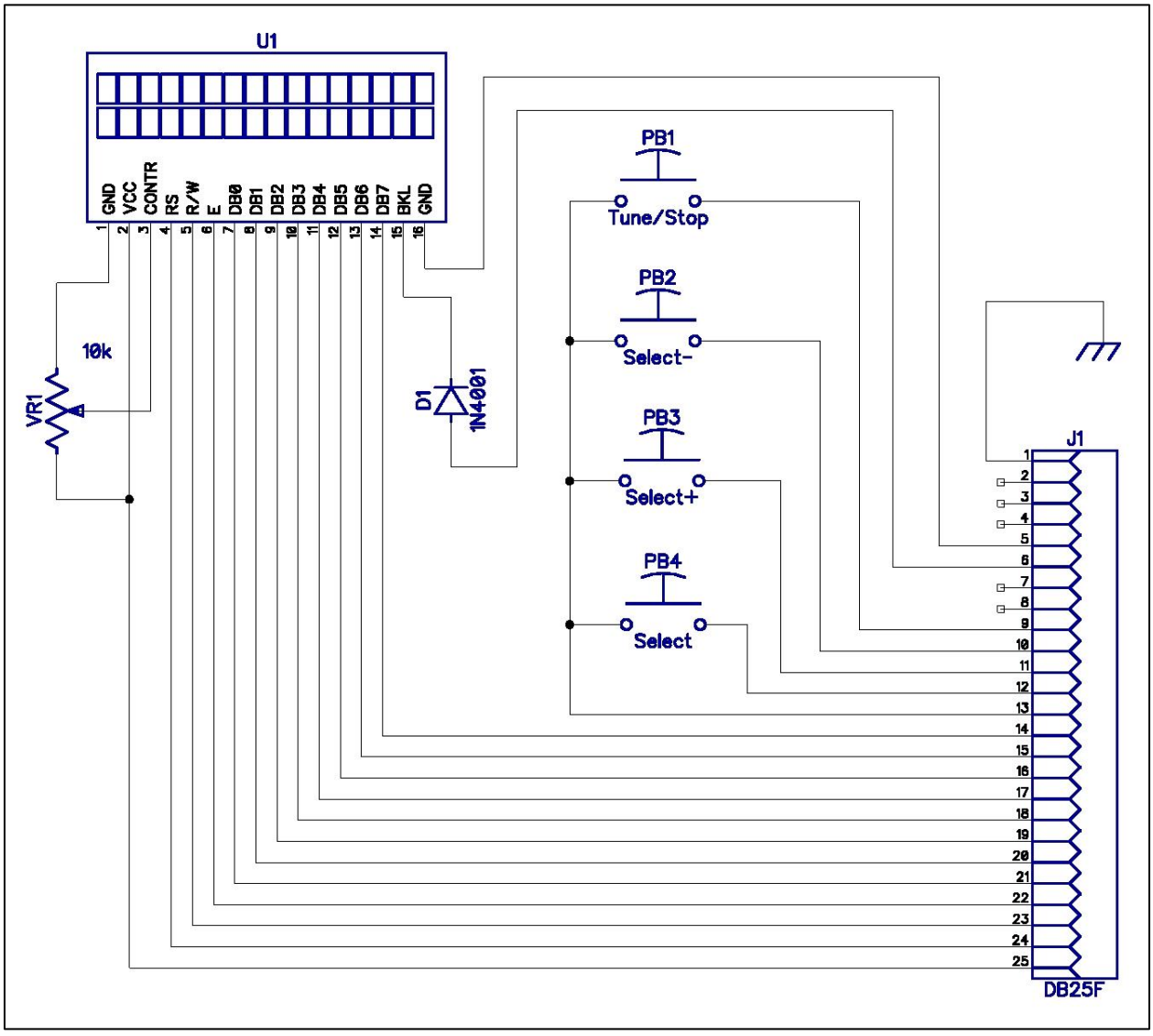
5. Directional Coupler



6. RF Power Sensors



7. Front Panel



Appendix E: Software

See Next Page

```
/*
* ---------------------------------------------------------------------
* PROGRAM DESCRIPTION
* ---------------------------------------------------------------------
* Company: Cal Poly State University
* Engineer: William Blodgett,
* Create Date: 6/7/2012 5:12:49 PM
* Description: This program runs in an ATmega32 MCU controlling an
* automatic antenna impedance matching system, also known as an
* antenna tuner. The matching network is series capacitor, shunt
* inductor L network optimized for 3.5 to 30 MHz operation. The
* inductor and capacitor are continuously variable and actuated
* by stepper motors. The shunt inductor is switchable from the load
* side to the source side providing wider impedance matching range.
* The switch is thrown by a third linear actuator stepper motor.
* The system uses a directional coupler and RF power
* sensors for forward and reflected power measurement. It displays
* this data, along with calculated SWR, inductor position, capacitor
* position on a 2 line by 16 character LCD. SWR = Standing Wave Ratio.
* It is a measure of impedance mismatch severity.
* Higher SWR = higher impedance mismatch. SWR of 1.0 is most desirable.
* The user initiates automatic tuning by first transmitting into
* the tuner and then pressing the Tune button.
* The system searches through the inductance and capacitance values
* to find the minimum SWR and improve impedance match. It attempts to
* reduce SWR to less than 1.5:1.
*
* ATUsysTest1.c
*
*/

// Define CPU Clock Frequency for delay functions
#define F_CPU 16000000

// EEPROM memory locations to store L, C, and Hi/Low Z Switch Settings
#define L_EEPROM 10
#define C_EEPROM 20
#define Z_EEPROM 30

// delay time in milliseconds for while(1) loop in main function
// sets display update rate and motor speed during manual control
#define MAIN_LOOP_DELAY 2


#define INDUCTOR_MAX 3500 // maximum inductor position

//C library files
#include <avr/io.h>       // AVR input/output library
#include <util/delay.h>   // software delay functions
#include <stdlib.h>       // contains itoa function
#include <stdint.h>       // AVR variable type definitions
#include <avr/eeprom.h>       // AVR EEPROM Read/Write Functions

/* Function Prototypes*/
void LCD_Initialization ();       // LCD Display power up sequence
void Write_Message ();            // writes ASCII strings to LCD
void clear_display ();            // clears the display
void Initialize_ADC0(void);       // Sets up ADC with required settings
void check_buttons(void);     // checks if button has been pressed
void set_write_location();        // sets character location on LCD
void update_display();            // updates data on display
uint32_t  isqrt32 (uint32_t n); // integer square root function
// samples fwd/refl power and calculates SWR
void calculate_SWR();
void auto_tune(void);             // performs automatic tuning
void tune_inductor(void);     // fine tunes inductor for minimum SWR
void tune_capacitor(void);        // fine tunes capacitor for minimum SWR
// returns inductor and capacitor back to start positions
void return_to_start(void);
void return_inductor_to_start();  // sets L to minimum
void return_capacitor_to_start(); // set C to maximum
void search_inductance();
void search_capacitance();

/* Strings of text for display */
char forward[3]  = "F";
char reflected[3]    = "R";
char SWRdisp[5]  = "SWR";
char d_point[3]      = ".";
char ascii_zero[3]   = "0";
char Ldisp[3]        = "L";
char Cdisp[3]        = "C";
char SLSC[6]     = "SLSC";
char SCSL[6]     = "SCSL";
```

```
85  char Select[3]       = ">";
86
87  // this Look Up Table (LUT) avoids need for floating point math
88  // by giving pre-calculated results of math function:
89  // (power, in watts) = [10^((power in dBm)/10)]/1000
90  // The power detectors output a DC voltage proportional to their
91  // input power in dBm.
92  // The values in this LUT are in units of Watts * 100.
93  // Each consecutive value is 0.1 dB change in power
94  volatile static uint16_t Power_LUT[471] =
95      {
96          1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 2, 2,
97          2, 2, 2, 2, 2, 2, 2, 2, 2, 2, 2, 2, 2, 2, 2, 2, 2, 2, 2, 2,
98          3, 3, 3, 3, 3, 3, 3, 3, 3, 3, 3, 3, 3, 3, 3, 4, 4, 4, 4, 4,
99          4, 4, 4, 4, 4, 4, 5, 5, 5, 5, 5, 5, 5, 5, 5, 6, 6, 6, 6, 6,
100         6, 6, 7, 7, 7, 7, 7, 7, 8, 8, 8, 8, 8, 9, 9, 9, 9, 9, 10,
101         10, 10, 10, 10, 11, 11, 11, 11, 12, 12, 12, 13, 13, 13, 13,
102         14, 14, 14, 15, 15, 15, 16, 16, 17, 17, 17, 18, 18, 19, 19,
103         19, 20, 20, 21, 21, 22, 22, 23, 23, 24, 25, 25, 26, 26, 27,
104         28, 28, 29, 30, 30, 31, 32, 32, 33, 34, 35, 35, 36, 37, 38,
105         39, 40, 41, 42, 43, 44, 45, 46, 47, 48, 49, 50, 51, 52, 54,
106         55, 56, 58, 59, 60, 62, 63, 65, 66, 68, 69, 71, 72, 74, 76,
107         78, 79, 81, 83, 85, 87, 89, 91, 93, 95, 98, 100, 102, 105,
108         107, 110, 112, 115, 117, 120, 123, 126, 129, 132, 135, 138,
109         141, 145, 148, 151, 155, 158, 162, 166, 170, 174, 178, 182,
110         186, 191, 195, 200, 204, 209, 214, 219, 224, 229, 234, 240,
111         245, 251, 257, 263, 269, 275, 282, 288, 295, 302, 309, 316,
112         324, 331, 339, 347, 355, 363, 372, 380, 389, 398, 407, 417,
113         427, 437, 447, 457, 468, 479, 490, 501, 513, 525, 537, 550,
114         562, 575, 589, 603, 617, 631, 646, 661, 676, 692, 708, 724,
115         741, 759, 776, 794, 813, 832, 851, 871, 891, 912, 933, 955,
116         977, 1000, 1023, 1047, 1072, 1096, 1122, 1148, 1175, 1202,
117         1230, 1259, 1288, 1318, 1349, 1380, 1413, 1445, 1479, 1514,
118         1549, 1585, 1622, 1660, 1698, 1738, 1778, 1820, 1862, 1905,
119         1950, 1995, 2042, 2089, 2138, 2188, 2239, 2291, 2344, 2399,
120         2455, 2512, 2570, 2630, 2692, 2754, 2818, 2884, 2951, 3020,
121         3090, 3162, 3236, 3311, 3388, 3467, 3548, 3631, 3715, 3802,
122         3890, 3981, 4074, 4169, 4266, 4365, 4467, 4571, 4677, 4786,
123         4898, 5012, 5129, 5248, 5370, 5495, 5623, 5754, 5888, 6026,
124         6166, 6310, 6457, 6607, 6761, 6918, 7079, 7244, 7413, 7586,
125         7762, 7943, 8128, 8318, 8511, 8710, 8913, 9120, 9333, 9550,
126         9772, 10000, 10233, 10471, 10715, 10965, 11220, 11482,
127         11749, 12023, 12303, 12589, 12882, 13183, 13490, 13804,
128         14125, 14454, 14791, 15136, 15488, 15849, 16218, 16596,
129         16982, 17378, 17783, 18197, 18621, 19055, 19498, 19953
130     };
131
132 /* Global Variables */
133 // function select: 1 = L, 2 = C, 3 = Switch, 0 = standby
134 volatile uint8_t function_select = 0;
135 int16_t capacitor = 1;                  // int value of capacitor position
136 int16_t inductor = 1;                   // int value of inductor position
137 uint8_t L_Network_Select = 1;     // 0 = SLSC Z, 1 = SCSL
138 uint16_t SWR = 0;                       //  int value of SWR
139 uint32_t Forward_Power    = 0;          //  int value of Forward Power
140 uint32_t Reflected_Power = 0;     //  int value of Reflected Power
141 uint8_t standby_count = 0;    //counter to time system standby mode entry
142 uint8_t tune_flag = 0;                  // true when auto tuning is enabled
143 // rts_flag = 1 if return_to_start hasn't been called
144 uint8_t rts_flag = 0;
145
146 int main(void)
147     {
148         // Data direction register port B set for input and output,
149         // 1=out 0=input
150         DDRB = 0b10000111;
151         DDRC = 0xff; // Data direction register port C set to output
152         DDRD = 0xff; // Data direction register port D set to output
153
154         // set bits 3 to 6 high for button pull up
155         PORTB = (PORTB | 0b01111000);
156         // initialize Port C to disable motors EN = 1 = disable
157         PORTC = 0b10010010;
158         // if motor is enabled, it draws full current ~1 AMP.
159         // Can't enable all 3 motors at once.
160         // since power supply limit is 2 amps
161
162         /* Local Variables */
163         // counter to time LCD update rate
164         uint8_t display_update_count = 0;
165
166         Initialize_ADC0();              // turn on ADC
167         LCD_Initialization ();          // Run LCD Initialization Function
168
```

```
169           // Read EEPROM stored values to recall positions at power off
170           // recall inductor position
171           inductor = eeprom_read_word((uint16_t *) L_EEPROM);
172           // recall capacitor position
173           capacitor = eeprom_read_word((uint16_t *) C_EEPROM);
174           // recall Switch Position
175           L_Network_Select = eeprom_read_byte((uint8_t *) Z_EEPROM);
176
177           // L_Network_Select can only be 1 or 0
178           if((L_Network_Select != 1)||(L_Network_Select != 0))
179               L_Network_Select = 1;
180
181           while(1)     // continuous loop
182               {
183                   // check for button press and perform button functions
184                   check_buttons();
185
186                   if(tune_flag == 1)    // if Tune button has been pressed
187                       auto_tune(); // run auto tuning function
188
189                   display_update_count++;
190                   // update display when display_update_count reaches 50
191                   if (display_update_count == 50)
192                       {
193                           calculate_SWR(); // get new SWR and power values
194                           update_display();     // write new values to LCD
195                           display_update_count = 0; // set count back to zero
196                           standby_count++;
197                      // standby time >
198                  // display_update_count * standby_count * MAIN_LOOP_DELAY, ms
199                       }
200
201                   // puts system in standby if no buttons are
202                   // pressed for standby time
203                   if(standby_count == 100)
204                  {
205                  function_select = 0; // put in standby mode
206                  PORTC = 0b10010010;  // turn off motors
207                  standby_count = 0; // reset standby counter
208                  // eeprom write functions placed here to
209                  // limit number of writes eeprom update
210                  // functions only update eeprom if stored
211                  // value doesn't match value to be written
212                  eeprom_update_word((uint16_t *) L_EEPROM, inductor);
213                  eeprom_update_word((uint16_t *) C_EEPROM, capacitor);
214                  eeprom_update_byte((uint8_t *) Z_EEPROM, L_Network_Select);
215                  }
216
217                   // this delay along with the display_update_count
218                   // control display refresh time
219                   // this delay also sets the motor speed during
220                   // manual control
221                   _delay_ms(MAIN_LOOP_DELAY);
222
223               }
224           return(0);
225       }
226
227       /* Function Definitions */
228
229       //LCD Initialization Function
230       void LCD_Initialization ()
231           {
232               _delay_ms(40);       // delay for LCD power on, minimum 30 ms
233               PORTB = (PORTB & 0b11111000);// Set Enable,RS,RW Bits to Low
234               _delay_us(1);                // Delay before Enable Bit High
235               PORTB = (PORTB | 0x04);      // Set Enable Bit to High
236               PORTD = 0x38;                // Function Set - Send 00 00111000
237               // Hold Enable Bit to High to allow data to arrive on port
238               _delay_us(2);
239               // Set Enable Bit to Low, data is read
240               PORTB = (PORTB & 0b11111000);
241               _delay_us(50);   // Delay for instruction process, minimum 39us
242               PORTB = (PORTB | 0x04);      // Set Enable Bit to High
243               PORTD = 0x0F;                // Display ON - Send   00 00001111
244               // Hold Enable Bit to High to allow data to arrive on port
245               _delay_us(2);
246               PORTB = (PORTB & 0b11111000);     // Set Enable Bit to Low
247               // Delay for instruction process, minimum 39us
248               _delay_us(50);
249               PORTB = (PORTB | 0x04);           // Set Enable Bit to High
250               PORTD = 0x01;            // Display Clear-Send  00 00000001
251               // Hold Enable Bit to High to allow data to arrive on port
252               _delay_us(2);
```

```
253                 PORTB = (PORTB & 0b11111000);     // Set Enable Bit to Low
254                 _delay_ms(2);         // Delay for Clear process, minimum 1.53ms
255                 PORTB = (PORTB | 0x04);      // Set Enable Bit to High
256                 PORTD = 0x06;                 // Entry Mode Set-Send 00 00000110
257                 // Hold Enable Bit to High to allow data to arrive on port
258                 _delay_us(2);
259                 //Set DDRAM Address to 0x00 to Display on First Line of LCD
260                 PORTB = (PORTB & 0b11111000);     // Set Enable Bit to Low
261                 _delay_us(50);                // Delay for instruction process
262                 PORTB = (PORTB | 0x04);      // Set Enable Bit to High
263                 // Set DDRAM Address to 0x00, top left of display
264                 PORTD = 0x80;
265                 // Hold Enable Bit to High to allow data to arrive on port
266                 _delay_us(2);
267                 PORTB = (PORTB & 0b11111000);     // Set Enable Bit to Low
268                 _delay_us(50);
269             }
270
271     // Function to write strings of ASCII characters to the LCD
272     void Write_Message (char* input)
273         {
274             char* ps = input;
275             while(*ps != '\0')
276                 {
277                     _delay_us(10);
278                     // Set Register Select and Enable Bit to High
279                     PORTB = (PORTB | 0x05);
280                     PORTD = *ps;    // Send ASCII character to input
281                     _delay_us(2);
282                     // Set Enable Bit to Low, keep Register Select High
283                     PORTB = (PORTB & 0b11111011);
284                     _delay_us(50);
285                     // Set Register Select low
286                     PORTB = (PORTB & 0b11111000);
287                     ps++;
288                 }
289         }
290
291     /* clears previously written data from LCD */
292     void clear_display ()
293         {
294             PORTB = (PORTB | 0x04);       // Set Enable Bit to High
295             PORTD = 0x01;             //Display Clear-Send  00 00000001
296             // Hold Enable Bit to High to allow data to arrive on port
297             _delay_us(2);
298             PORTB = (PORTB & 0b11111000); // Set Enable Bit to Low
299             _delay_ms(2);         // Delay for Clear process, minimum 1.53ms
300             PORTB = (PORTB | 0x04);       // Set Enable Bit to High
301             PORTD = 0x80;// Set DDRAM Address to 0x00, top left of display
302            // 0x80 because top bit needs to be "1" to set DDRAM address
303             // Hold Enable Bit to High to allow data to arrive on port
304             _delay_us(2);
305             PORTB = (PORTB & 0b11111000); // Set Enable Bit to Low
306             _delay_us(50);
307         }
308
309          // this function sets up ADC with required settings
310          void Initialize_ADC0(void)
311         {
312             //MAX A/D conversion rate = (ADC Clock)/(13 clock cycles)
313             // ADC Clock Freq = (CPU Clk Freq) / (prescaler value)
314             ADCSRA = 0x87;//Turn On ADC and set prescaler (CLK/128=125 kHz)
315             // Set Ref Sel for 2.56V ref, right adjust result,
316             // select ADC channel 0
317             ADMUX = 0xC0;
318             // ADMUX = 0xC1 for selecting ADC channel 1
319             _delay_us(260);       //25 cycles or 260us (max for start-up)
320         }
321
322     /* Sets Location on LCD where data will be written */
323     void set_write_location(int input)
324         {
325             PORTB = (PORTB | 0x04);       // Set Enable Bit to High
326             PORTD = input;                    // Set DDRAM Address
327             // Hold Enable Bit to High to allow data to arrive on port
328             _delay_us(2);
329             PORTB = (PORTB & 0b11111000);     // Set Enable Bit to Low
330             _delay_us(50);
331         }
332
333     /* This function checks for a button press
334     and modifies current operating mode */
335     void check_buttons(void)
336         {
```

```
337             int i;          // temporary variable used in FOR loops
338             if (!(PINB & 0x08))       // if Select button is pressed
339                 {
340                     _delay_ms(250);          // delay for button press
341                     function_select++;       // increment function_select
342                     // Reset all Motor Enable Bits to 1, all motors off
343                     PORTC = (PORTC | 0b10010010);
344                     _delay_ms(250);              // delay for button press
345                     if (function_select == 1)     // L selected
346                         PORTC &= ~(1 << 1);  // Set PC1 low, turns on L Motor
347                     if (function_select == 2)     // C selected
348                         // Set PC7 low to turn on C Motor
349                         PORTC &= ~(1 << 7);
350                     if (function_select == 3)     // Hi/Low Z selected
351                         // Set PC4 low to turn on Switch Motor
352                         PORTC &= ~(1 << 4);
353                     if (function_select == 4)     // 4 is not a valid function
354                         // roll function select back to zero
355                         function_select = 0;
356                     if (function_select == 0)
357                         PORTC = (PORTC | 0b10010010);// Turn off All Motors
358                     standby_count = 0;   // button press, reset standby count
359                 }
360
361             if (!(PINB & 0x10))           // if Select + button is pressed
362                 {
363                     standby_count = 0;   // button pressed, reset standby count
364                     if (function_select == 1)     // L selected
365                         {
366                             // L motor driver set to full step
367                             // (200 steps/revolution)
368                             // Lmin = 0, Lmax = 8500
369                             // set DIR high for increasing direction
370                             PORTB |= (1 << 7);
371                             PORTC |= (1 << 0);       // make Step go high
372                             _delay_us(10);           // delay for driver
373                             PORTC &= ~(1 << 0);      // make Step go low
374                             _delay_us(10);           // delay for driver
375                             inductor++;              // increment position
376                         }
377
378                     if (function_select == 2)         // C selected
379                         {
380                             // C Stepper Driver Set to 16th Step
381                             // Cmin = 0, Cmax = 1660
382                             _delay_ms(10);           // delay for motor
383                             // set DIR low for increasing direction
384                             PORTC &= ~(1 << 5);
385                             PORTC |= (1 << 6);   // set Step high
386                             _delay_us(10);       // delay for motor driver
387                             PORTC &= ~(1 << 6);  // set Step low
388                             _delay_us(10);       // delay for motor driver
389                             capacitor++; // increment capacitor position
390                         }
391                     // L network switch position selected
392                     if (function_select == 3)
393                         {
394                             if (L_Network_Select == 0)
395                                 {
396                                     // delay for button press
397                                     _delay_ms(250);
398                                     // toggle function
399                                     L_Network_Select = 1;
400                                     // set DIR low for increasing direction
401                                     PORTC &= ~(1 << 2);
402                                     // i counts from 0 to 899, number of
403                                     // steps required to throw switch
404                                     for (i= 0; i< 900; i++)
405                                         {
406                                             // delay for step speed
407                                             _delay_ms(10);
408                                             // set Step high
409                                             PORTC |= (1 << 3);
410                                             // delay for motor driver
411                                             _delay_us(10);
412                                             // set Step low
413                                             PORTC &= ~(1 << 3);
414                                             // delay for motor driver
415                                             _delay_us(10);
416                                         }
417                                     // delay for button press
418                                     _delay_ms(250);
419                                 }
420
```

```
                        else if (L_Network_Select == 1)
                            {
                                // delay for button press
                                _delay_ms(250);
                                // toggle function
                                L_Network_Select = 0;
                                // set DIR high for decreasing direction
                                PORTC |= (1 << 2);
                                // i counts from 0 to 899
                                for (i= 0; i< 900; i++)
                                    {
                                        // delay for step speed
                                        _delay_ms(10);
                                        // set Step high
                                        PORTC |= (1 << 3);
                                        // delay for driver
                                        _delay_us(10);
                                        // set Step low
                                        PORTC &= ~(1 << 3);
                                        // delay for driver
                                        _delay_us(10);
                                    }
                                // delay for button press
                                _delay_ms(250);
                            }
                    }
            }

        if (!(PINB & 0x20)) // if Select- button is pressed
            {
                // button pressed, reset standby count
                standby_count = 0;
                if (function_select == 1)     // L selected
                    {
                        // Lmin = 0, Lmax = 8500 steps
                        // set DIR low for decreasing direction
                        PORTB &= ~(1 << 7);
                        PORTC |= (1 << 0);        // set Step high
                        _delay_us(10);        // delay for driver
                        PORTC &= ~(1 << 0);  // set Step low
                        _delay_us(10);        // delay for driver
                        inductor--;           // decrement position
                    }

                if (function_select == 2)     // C selected
                    {
                        // C Stepper Driver Set to 16th Steps
                        // 3200 steps per revolution
                        // Cmin = 0, Cmax = 1660
                        _delay_ms(10);        // delay for button press
                        // set DIR high for decreasing direction
                        PORTC |= (1 << 5);
                        PORTC |= (1 << 6);   // set Step high
                        _delay_us(10);        // delay for motor driver
                        PORTC &= ~(1 << 6);  // set Step low
                        _delay_us(10);        // delay for motor driver
                        capacitor--;     // decrement position
                    }
            }

        if (!(PINB & 0x40)) // if Tune/Stop button is pressed
            {
                _delay_ms(500);       // delay for button press
                tune_flag = 1;   // indicates automatic tuning selected
                rts_flag = 1;
                _delay_ms(500);
            }
    }

// updates data on LCD
void update_display()
    {
        // divide power values by 100 for power in Watts
        Forward_Power = (Forward_Power / 100);
        Reflected_Power = (Reflected_Power / 100);
        // calculate whole number part of SWR
        uint8_t SWR_whole = SWR/100;
        // calculate decimal part of SWR
        uint8_t SWR_decimal = SWR%100;
        //  string for displayed ascii digits
        char disp_ascii[20];
        clear_display ();
        Write_Message (forward);      // Display "F"
        // convert int to ASCII string
```

```
505             itoa(Forward_Power, disp_ascii, 0xA);
506             Write_Message (disp_ascii);       // Display forward power value
507             // set display location for reflected power
508             set_write_location(0x85);
509             Write_Message (reflected);        // Display "R"
510             // convert int to ASCII string
511             itoa (Reflected_Power, disp_ascii,  0xA);
512             Write_Message (disp_ascii);   // Display reflected power value
513             set_write_location(0x89); // set display location for SWR
514             Write_Message (SWRdisp);  // Display "SWR"
515             // convert int to ASCII string
516             itoa (SWR_whole, disp_ascii, 0xA);
517             Write_Message (disp_ascii);   // Display SWR whole number value
518             Write_Message(d_point);       // place decimal point
519             if (SWR_decimal < 10)    // if the decimal value is less than 10
520                 Write_Message(ascii_zero);    // place a zero in front of it
521             itoa(SWR_decimal, disp_ascii, 0xA);
522             Write_Message(disp_ascii);
523             if (function_select == 1)             // L selected
524                 {
525                     set_write_location(0xC0); // place select symbol at L
526                     Write_Message (Select);
527                 }
528             set_write_location(0xC1);     // set display location for L
529             Write_Message (Ldisp);            // Display "L"
530             itoa (inductor, disp_ascii, 0xA);
531             Write_Message (disp_ascii);       // Display inductor value
532             if (function_select == 2)         // C selected
533                 {
534                     set_write_location(0xC6); // place select symbol at C
535                     Write_Message (Select);
536                 }
537             set_write_location(0xC7);     // set display location for C
538             Write_Message (Cdisp);            // Display "C"
539             itoa (capacitor, disp_ascii, 0xA);
540             Write_Message (disp_ascii);       // Display capacitor value
541             if (function_select == 3)         // Switch selected
542                 {
543                     // place select symbol at SLSC/SCSL
544                     set_write_location(0xCB);
545                     Write_Message (Select);
546                 }
547             set_write_location(0xCC);     // set display location
548             if (L_Network_Select == 0)
549                 Write_Message (SLSC);         // Display "SLSC"
550             if (L_Network_Select == 1)
551                 Write_Message (SCSL);         // Display "SCSL"
552         }
553
554     // this function gets data from the ADC and produces SWR,
555     // Forward_Power, and Reflected_Power global
556     // variable results the SWR, Forward_Power, and
557     // Reflected_Power results are 100 times higher to
558     // preserve fractional part. For example, if the
559     // Forward_Power result is 1020, the actual value in
560     // watts is 1000/100 or 10.2 watts the analog voltage
561     // converted by the ADC is the output of the
562     // forward and reflected power detectors whose DC output
563     // voltage is linearly proportional to their
564     // input power in dBm
565     void calculate_SWR()
566         {
567             uint32_t Reflected_Power_numer;   // numerator of SWR Fraction
568             uint32_t SWR_fraction;    // result of fraction used in SWR calc
569             uint32_t SWR_root;    // result of square root used in SWR calc
570             uint16_t Power_index;// stores ADC result and indexes Power_LUT
571             ADMUX = 0xC0;     // Select ADC0 to sample forward power
572             ADCSRA = 0xC7;    // Start ADC conversion
573             _delay_us(120); // wait for ADC conversion
574             Power_index = ADC;    // assigns ADC converted value to variable
575             // this makes any power less than offset = 0 Watts
576             if (Power_index < 495)
577                 Power_index = 0;
578             else
579                 // allow indexing of proper value in LUT
580                 Power_index = Power_index - 495;
581                 // if indicated power is reading lower than actual power
582                 // then:
583                 // decrease value subtracted from power index to correct
584                 // each incremental value of the Power_index is
585                 // approximately equivalent to 0.1 dB
586
587             Forward_Power = Power_LUT[Power_index]; // assigns value in LUT to
588             // forward power variable
```

```
589
590             ADMUX = 0xC1;    // Select ADC1
591             ADCSRA = 0xC7;   // Start ADC conversion
592             _delay_us(120); // wait for ADC conversion
593             Power_index = ADC; // get value from ADC
594             if (Power_index < 506) // any power less than offset = 0 Watts
595                 Power_index = 0;
596             else
597                 Power_index = Power_index - 506;
598
599             Reflected_Power = Power_LUT[Power_index]; // get reflected
600             // power value from LUT
601
602             /*
603             SWR Calculation performs:
604             SWR =
605             [1 + sqrt((refl pwr)/(fwd pwr))]/[1 - sqrt((refl pwr)/(fwd pwr))]
606             */
607             // can only calculate valid SWR if transmit power > 3 Watts
608             if (Forward_Power > 300)
609                 {
610                      // multiply up to preserve accuracy before division
611                     Reflected_Power_numer = Reflected_Power * 10000;
612                     SWR_fraction = Reflected_Power_numer/Forward_Power;
613                     SWR_root = isqrt32 (SWR_fraction);
614                     SWR = (10000 + (100 * SWR_root))/(100 - SWR_root);
615                      // if reflected pwr > fwd pwr, SWR result negative
616                     if(SWR < 100)
617                         // reset negative values back to positive, very hi SWR
618                         SWR = 10000;
619                 }
620         }
621
622     // This function performs the automatic impedance matching routine.
623     void auto_tune(void)
624     {
625         // tune_count counts number of L then C tuning iterations
626         int tune_count = 0;
627         // run calculate SWR function to determine if transmit power is applied
628
629         calculate_SWR();
630
631         // the code in the following IF statement performs an initial search
632         // through entire range of inductance
633         // and capacitance to find settings for minimum SWR
634         // only proceeds if transmitter input power is greater than 3 W
635         // and SWR is high and rts_flag ensures
636         // only one search through full range of L and C
637         if (Forward_Power>300 && !(SWR >=100 && SWR < 150 ) && rts_flag == 1)
638             {
639                 return_inductor_to_start();       // sets L to minimum
640                 return_capacitor_to_start();  // set C to maximum
641                 search_inductance();
642                // next search capacitance range for minimum SWR
643                 search_capacitance();
644
645                // set rts_flag to zero to ensure that entire
646                // range of L and C only searched once
647                 rts_flag = 0;
648             }
649
650         // set DIR high for decreasing capacitance direction
651         PORTC |= (1 << 5);
652
653         // the code in the following IF statement performs fine tuning of
654         // L and C to get SWR < 1.5
655         // only tune if transmitter input power is greater than
656         // 5 W and SWR is not between 1.0 and 1.5
657         if (Forward_Power > 500 && !(SWR >=100 && SWR < 150 ))
658             {
659                  // tune_flag set by button press, allows manual start/stop
660                 // tune_count set for 4 tuning attempts
661                 while (tune_flag == 1 && tune_count < 4)
662                     {
663                         calculate_SWR();
664                         // if SWR is not between 1.0 and 1.5
665                         if(!(SWR >=100 && SWR < 150 ))
666                             {
667                                 // tune inductor
668                                 // Turn off All Motors
669                                 PORTC = (PORTC | 0b10010010);
670                                 // Set PC1 low to turn on L Motor
671                                 PORTC &= ~(1 << 1);
672                                 tune_inductor(); // tune inductance
```

```
673
674                                 // tune capacitor
675                                 // Turn off All Motors
676                                 PORTC = (PORTC | 0b10010010);
677                                 // Set PC7 low to turn on C Motor
678                                 PORTC &= ~(1 << 7);
679                                 tune_capacitor();     // tune capacitance
680                             }
681                         tune_count++;    // increment tune count
682                     }
683             }
684         else
685             // set the tune_flag to zero to stop tuning if no RF power applied
686             tune_flag = 0;
687         PORTC = (PORTC | 0b10010010);         // Turn off All Motors
688         tune_flag = 0;   // set the tune_flag to zero to stop tuning function
689     }
690
691     // performs fine tuning of inductor
692     void tune_inductor(void)
693     {
694         uint16_t i = 0;                             // FOR loop count
695         // counts number of steps for which SWR increases
696         int8_t increasing_SWR_count = 0;
697         uint8_t toggle_count = 0;                   // counts direction changes
698         uint16_t minimum_value = SWR;          // stores minimum SWR value
699         int16_t inductor_minimum2 = inductor;// stores location of minimum SWR
700
701         do
702         {
703             _delay_us(10);            // delay for motor driver
704             PORTC |= (1 << 0);        // set Step high
705             _delay_us(10);            // delay for driver
706             PORTC &= ~(1 << 0);       // set Step low
707             _delay_us(10);            // delay for driver
708             if (PORTB & (1 << 7))     // if increasing direction selected
709                 inductor++;       // increment displayed rel. inductance value
710             else
711                 inductor--;       // decrement displayed rel. inductance value
712             _delay_ms(20);        //  delay for stepper motor speed limitation
713             //_delay_us(500);         // added delay to fine tune motor speed
714             calculate_SWR();      // get new power & SWR values
715             if(inductor == 1)         // prevents passing mechanical stop
716                 PORTB ^= (1 << 7);    // toggle DIR bit to change direction
717             if (!(PINB & 0x40))       // if Tune/Stop button is pressed
718                 {
719                     _delay_ms(500); // delay for button de-bounce
720                     // set tune flag to false to stop auto tuning
721                     tune_flag = 0;
722                 }
723
724             if(SWR < minimum_value)   // if new SWR < stored minimum_value
725                 {
726                     minimum_value = SWR; // set new SWR value as the minimum
727                     // save the inductor position for minimum
728                     inductor_minimum2 = inductor;
729                 }
730             else
731                 increasing_SWR_count++;
732
733             // if increasing_SWR_count reaches limit or
734             // the mechanical limits are reached
735     if((increasing_SWR_count>110 )||( inductor < 5)||(inductor > INDUCTOR_MAX))
736             {
737                 PORTB ^= (1 << 7);        // toggle DIR bit to change direction
738                 toggle_count++;           // count # times direction changes
739                 increasing_SWR_count = 0;     // reset count
740             }
741         // only 2 direction changes and tuning hasn't been manually disabled
742         } while (toggle_count < 2 && tune_flag == 1);
743
744         // if current position higher than minimum SWR position
745         if(inductor > inductor_minimum2)
746             {
747                 // find # of steps to SWR minimum
748                 inductor_minimum2 = inductor - inductor_minimum2;
749                 // set DIR low for decreasing direction
750                 PORTB &= ~(1 << 7);
751                 // set inductor value to SWR minimum location
752                 inductor = inductor - inductor_minimum2;
753             }
754
755         else // if current position lower than minimum SWR position
756             {
```

```
757                 // find # of steps to SWR minimum
758                 inductor_minimum2 = inductor_minimum2  - inductor;
759                 // set DIR high for increasing direction
760                 PORTB |= (1 << 7);
761                 // set inductor value to SWR minimum location
762                 inductor = inductor + inductor_minimum2;
763             }
764
765         _delay_us(10);            // delay for driver
766         // FOR loop returns inductor to location of SWR minimum
767         for (i = 0; i < inductor_minimum2; i++)
768             {
769                 PORTC |= (1 << 0);        // set Step high
770                 _delay_us(10);            // delay for driver
771                 PORTC &= ~(1 << 0);       // set Step low
772                 _delay_ms(3);         // delay for stepper motor speed limit
773             }
774
775         //  if the SWR is still high, move inductor past minimum SWR
776         // point to get closer to R = 1 circle
777         if (SWR > 300)
778         {
779             for (i = 0; i < 51; i++)
780             {
781                 PORTB |= (1 << 7);    // set DIR high for increasing direction
782                 PORTC |= (1 << 0);        // set Step high
783                 _delay_us(10);            // delay for driver
784                 PORTC &= ~(1 << 0);       // set Step low
785                 _delay_ms(3);         //  delay for stepper motor speed limit
786                 inductor++;
787             }
788         }
789     }
790
791     // performs fine tuning of capacitor
792     void tune_capacitor(void)
793     {
794         uint16_t i = 0;           // FOR loop count
795         // counts number of steps for which SWR increases
796         uint8_t increasing_SWR_count = 0;
797         uint8_t toggle_count = 0;  // counts direction changes
798         // stores minimum SWR value for comparison to updated SWR
799         uint16_t minimum_value = SWR;
800         // stores location of minimum SWR
801         int16_t capacitor_minimum = capacitor;
802
803         do
804         {
805             _delay_us(10);            // delay for motor driver
806             PORTC |= (1 << 6);        // set Step high
807             _delay_us(10);            // delay for driver
808             PORTC &= ~(1 << 6);       // set Step low
809             _delay_us(10);            // delay for driver
810             if (PORTC & (1 << 5))     // if decreasing direction selected
811                 capacitor--;      // decrement  rel. capacitance value
812             else
813                 capacitor++; // increment capacitor position
814             _delay_ms(16);        // delay for motor speed limit
815             calculate_SWR(); // get new power & SWR values
816             if (!(PINB & 0x40)) // if Tune/Stop button is pressed
817                 {
818                     _delay_ms(500); // delay for button de-bounce
819                     // set tune flag to false to stop auto tuning
820                     tune_flag = 0;
821                 }
822             // if new SWR value less than stored minimum_value
823             if(SWR < minimum_value)
824                 {
825                     minimum_value = SWR;// set new SWR value as the minimum
826                     // save the capacitor position for minimum
827                     capacitor_minimum = capacitor;
828                 }
829             else
830                 increasing_SWR_count++;
831
832             // if increasing_SWR_count reaches limit
833             if(increasing_SWR_count > 250)
834             {
835                 PORTC ^= (1 << 5);        // toggle DIR bit to change direction
836                 toggle_count++;       // count how many times direction changes
837                 increasing_SWR_count = 0;     // reset count
838             }
839         } while (toggle_count < 2 && tune_flag == 1);
840
```

```
841         // if current position higher than minimum SWR position
842         if(capacitor > capacitor_minimum)
843             {
844                 // find # of steps to SWR minimum
845                 capacitor_minimum = capacitor - capacitor_minimum;
846                 PORTC |= (1 << 5);   // set DIR high for decreasing direction
847                 // set position to SWR minimum location
848                 capacitor = capacitor - capacitor_minimum;
849             }
850         else // if current position lower than minimum SWR position
851             {
852                 // find # of steps to SWR minimum
853                 capacitor_minimum = capacitor_minimum - capacitor;
854                 PORTC &= ~(1 << 5);  // set DIR low for increasing direction
855                 // set position to SWR minimum location
856                 capacitor = capacitor + capacitor_minimum;
857             }
858         _delay_us(10);            // delay for motor driver
859         // FOR loop returns capacitor to location of SWR minimum
860         for (i = 0; i < capacitor_minimum; i++)
861             {
862                 PORTC |= (1 << 6);       // set Step high
863                 _delay_us(10);           // delay for driver
864                 PORTC &= ~(1 << 6);      // set Step low
865                 _delay_ms(2);    // delay for stepper motor speed limitation
866             }
867     }
868
869     //Sets inductor to minimum value and capacitor to
870     // maximum value for auto tuning start point
871     void return_inductor_to_start(void)
872     {
873         int i; // temporary variable for FOR loops
874         // int inductor2 = inductor  + 1;
875         PORTC = (PORTC | 0b10010010);         // Turn off All Motors
876         _delay_us(10);            // delay for motor driver
877         PORTC &= ~(1 << 1);       // Set PC1 low to turn on L Motor
878         _delay_us(10);            // delay for motor driver
879         PORTB &= ~(1 << 7);       // set DIR low for decreasing direction
880         _delay_us(10);            // delay for motor driver
881         for (i = 0; i < inductor; i++)
882         {
883             PORTC |= (1 << 0);       // set Step high
884             _delay_us(10);           // delay for driver
885             PORTC &= ~(1 << 0);      // set Step low
886             _delay_ms(2);        // delay for stepper motor speed limitation
887         }
888         inductor = 0;                // reset displayed rel. inductance value
889
890         PORTC = (PORTC | 0b10010010);         // Turn off All Motors
891
892         _delay_us(10);            // delay for motor driver
893
894     }
895
896     void return_capacitor_to_start()
897     {
898         uint16_t i = 0;      // FOR loop counter
899
900         // 1660 = max capacitor position
901         int16_t capacitor_minimum = 1660 - capacitor;
902
903         PORTC &= ~(1 << 7);       // Set PC7 low to turn on C Motor
904         _delay_us(10);            // delay for motor driver
905         PORTC &= ~(1 << 5);       // set DIR low for increasing direction
906
907         // set capacitor to maximum C position
908         for (i = 0; i < capacitor_minimum; i++)
909         {
910             _delay_us(10);            // delay for motor driver
911             PORTC |= (1 << 6);        // set Step high
912             _delay_us(10);            // delay for driver
913             PORTC &= ~(1 << 6);       // set Step low
914             _delay_ms(2);             // delay for motor speed limit
915             capacitor++;          // increment capacitor position
916         }
917
918         PORTC = (PORTC | 0b10010010);         // Turn off All Motors
919     }
920
921
922     void search_inductance()
923     {
924         uint16_t i = 0;      // FOR loop counter
```

```
925
926         // run calculate SWR function to determine if
927         // transmit power is applied
928         calculate_SWR();
929         // minimum_value stores the minimum SWR value
930         uint16_t minimum_value = SWR;
931         // stores the inductor position for minimum SWR
932         int16_t inductor_minimum1 = inductor;
933
934         PORTC &= ~(1 << 1);       // Set PC1 low to turn on L Motor
935         _delay_us(10);            // delay for motor driver
936         PORTB |= (1 << 7);        // set DIR high for increasing direction
937         for (i=0; i<(INDUCTOR_MAX +1); i++)   // search  all inductor steps
938             {
939                 _delay_us(10);            // delay for motor driver
940                 PORTC |= (1 << 0);        // set Step high
941                 _delay_us(10);            // delay for motor driver
942                 PORTC &= ~(1 << 0);       // set Step low
943                 _delay_us(10);            // delay for motor driver
944                 inductor++;               // increment inductance position
945                 _delay_ms(3);         // delay for stepper motor speed limit
946                 calculate_SWR();      // get new power & SWR values
947                 // if new SWR value less than stored minimum_value
948                 if(SWR < minimum_value)
949                     {
950                         minimum_value = SWR; // set new SWR value as minimum
951                         // save the inductor position for minimum
952                         inductor_minimum1 = inductor;
953                     }
954             }
955         // find # of steps to SWR minimum
956         inductor_minimum1 = inductor - inductor_minimum1;
957         PORTB &= ~(1 << 7);       // set DIR low for decreasing direction
958         // set inductor value to SWR minimum location
959         inductor = inductor - inductor_minimum1;
960
961         _delay_us(10);            // delay for driver
962         // this FOR loop returns inductor to location of SWR minimum
963         for (i = 0; i < inductor_minimum1; i++)
964             {
965                 PORTC |= (1 << 0);        // set Step high
966                 _delay_us(10);            // delay for motor driver
967                 PORTC &= ~(1 << 0);       // set Step low
968                 _delay_ms(2);    //  delay for stepper motor speed limitation
969             }
970
971         PORTC = (PORTC | 0b10010010);         // Turn off All Motors
972     }
973
974     void search_capacitance()
975     {
976         uint16_t i = 0;       // FOR loop counter
977         // stores the capacitor position for minimum SWR
978         int16_t capacitor_minimum = capacitor;
979         // run calculate SWR function to determine if transmit power applied
980         calculate_SWR();
981         // minimum_value stores the minimum SWR value
982         uint16_t minimum_value = SWR;
983
984         _delay_us(10);                    // delay for motor driver
985                 PORTC &= ~(1 << 7);       // Set PC7 low to turn on C Motor
986                 _delay_us(10);            // delay for motor driver
987                 PORTC |= (1 << 5);   // set DIR high for decreasing direction
988
989                 // this FOR loop searches entire range of capacitance for
990                 // minimum SWR capacitance
991                 for (i = 0; i < 1660; i++)
992                     {
993                         _delay_us(10);            // delay for motor driver
994                         PORTC |= (1 << 6);        // set Step high
995                         _delay_us(10);            // delay for driver
996                         PORTC &= ~(1 << 6);       // set Step low
997                         // delay for motor speed limit & SWR measurement
998                         _delay_ms(10);
999                         capacitor--;      // decrement capacitor position
1000                        calculate_SWR();      // get new SWR value
1001                        // if new SWR value less than stored minimum_value
1002                        if(SWR < minimum_value)
1003                            {
1004                                // set new SWR value as the minimum
1005                                minimum_value = SWR;
1006                                // save the capacitor position for minimum
1007                                capacitor_minimum = capacitor;
1008                            }
```

```
                }
            // find # of steps to SWR minimum
             capacitor_minimum = capacitor_minimum - capacitor;
             PORTC &= ~(1 << 5);  // set DIR low for increasing direction
             // set the capacitor position to the minimum location
             capacitor = capacitor + capacitor_minimum;

             _delay_us(10);           // delay for motor driver
             // FOR loop returns capacitor to location of SWR minimum
             // return the capacitor to the SWR minimum position
             for (i = 0; i < capacitor_minimum; i++)
                 {
                     PORTC |= (1 << 6);       // set Step high
                     _delay_us(10);           // delay for motor driver
                     PORTC &= ~(1 << 6);      // set Step low
                     _delay_ms(2);// delay for stepper motor speed limit
                 }
             PORTC = (PORTC | 0b10010010);         // Turn off All Motors
}

// integer square root function takes square root of integer
// input and returns an integer
// from: http://www.codecodex.com/wiki/Calculate_an_integer_square_root
uint32_t  isqrt32 (uint32_t n)
    {
         register uint32_t root, remainder, place;
         root = 0;
         remainder = n;
         place = 0x40000000;
         while (place > remainder)
        place = place >> 2;
         while (place)
             {
                 if (remainder >= root + place)
                     {
                         remainder = remainder - root - place;
                         root = root + (place << 1);
                     }
                 root = root >> 1;
                 place = place >> 2;
             }
         return root;
    }
```

A Crystal Set for the Boy Builder

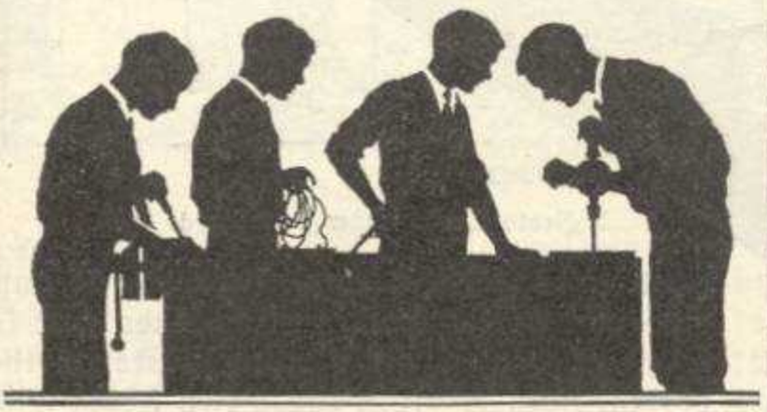
By WILL H. BATES

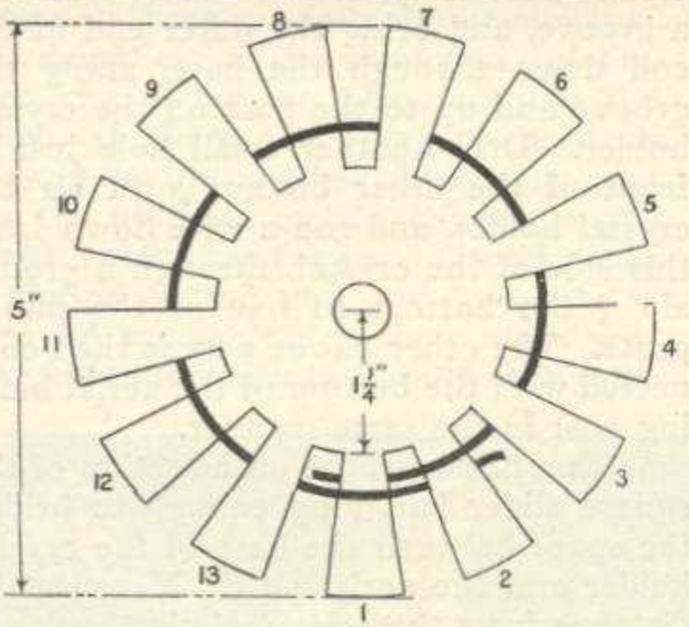
THIS little set is recommended for the boy builder, although the grown-ups also will enjoy it. Exclusive of the phones, the cost should not exceed $1. Surprising volume is obtained on the local stations, and the set is more selective than the average receiver of this type.

Very few parts are required. The wood base may be made of scrap material found around the home. The other items can be bought at the 10-cent store, with the exception of the crystal holder. For this, an upright type, as illustrated, should be selected, and may be obtained for about 25 cents. A good pair of 2,000-ohm phones should be used and these are available for about $2.50.

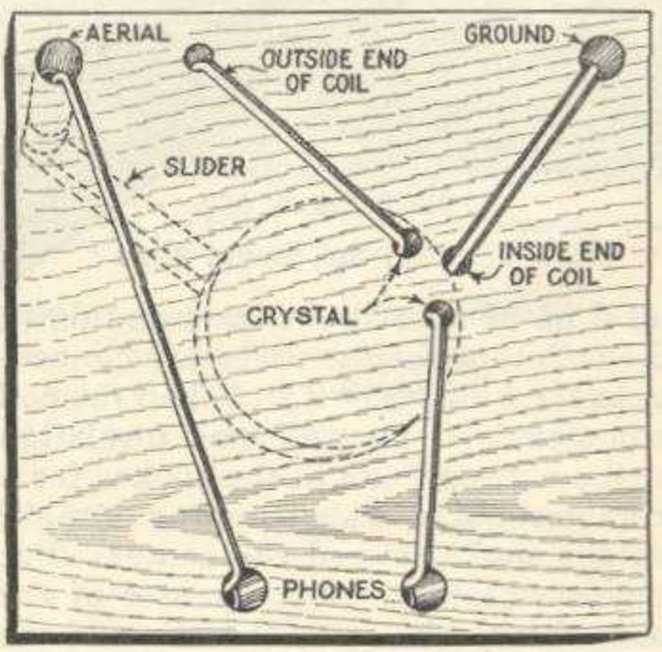
The base may be beveled on the top edges, sanded down, and then given a coat of mahogany or walnut stain. Drill it for the four binding posts, which are located as shown in the illustration of the under side of the base. A ½-in. bit is used to countersink these holes halfway through the board from below.

The spiderweb coil is wound on the form shown in the diagram; this form may be cut from fiber or cardboard, and the dimensions should be followed. The thirteen spokes are 1¼ in. long, leaving an uncut center 2½ in. in diameter. Wind the turns of No. 24 d.c.c. magnet wire alter-

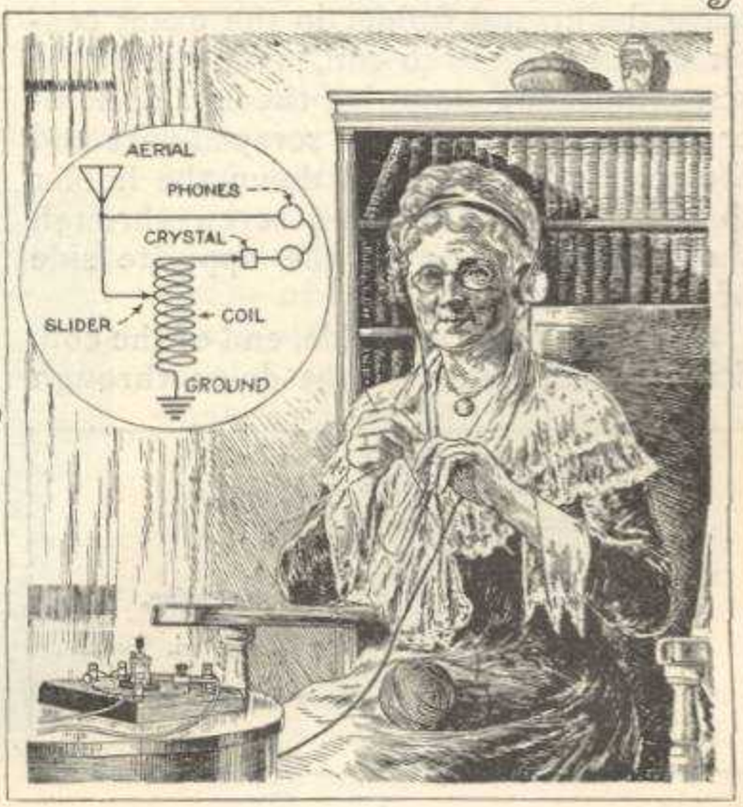




Left, Construction Details of Coil Form; Black Line Shows Method of Winding; Insert at Right, Schematic Circuit Diagram



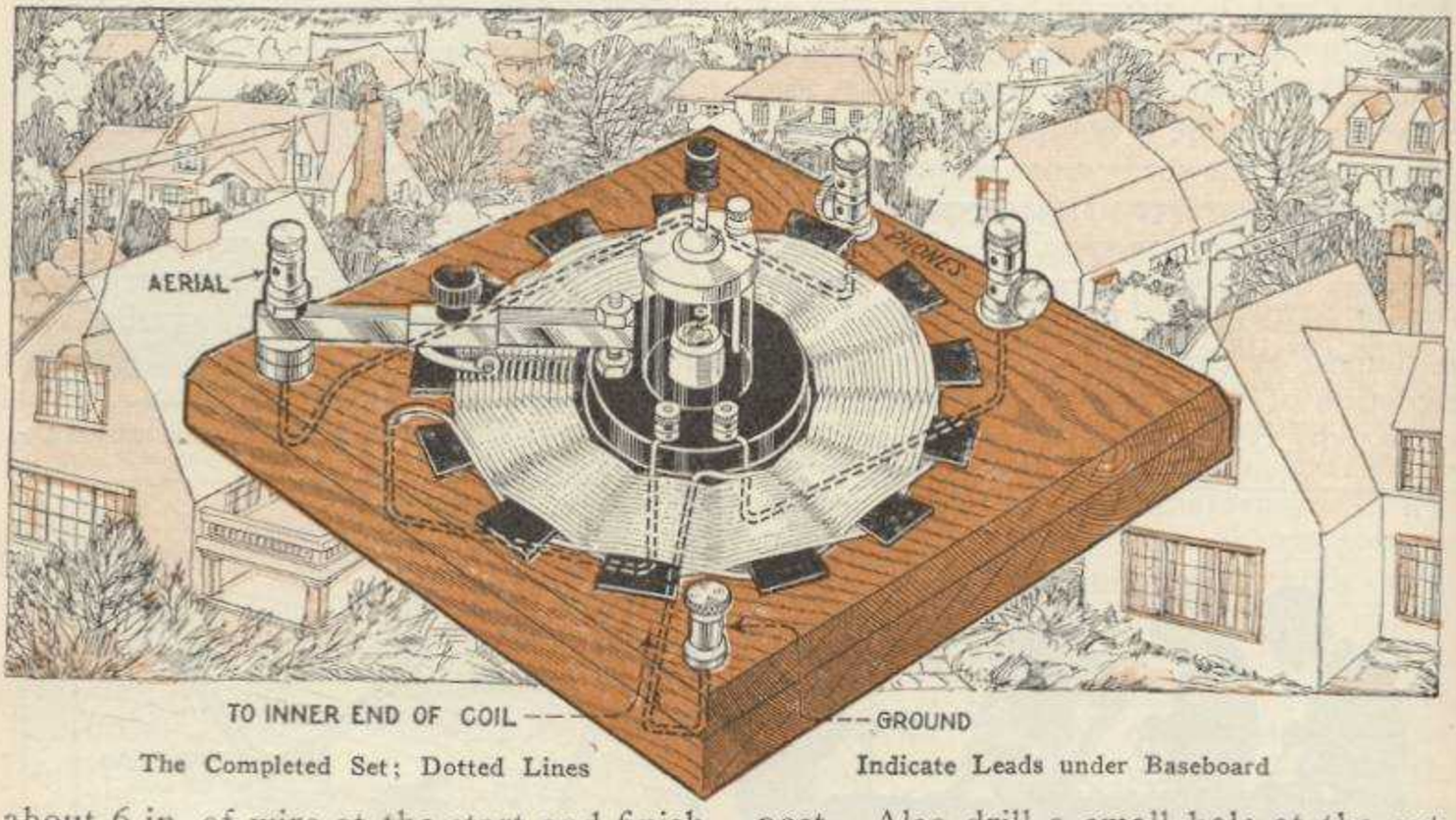
WIRING ON UNDER SIDE OF BASE



nately over and under the spokes, as indicated by the heavy line. When doing this keep count of the turns by marking one spoke and counting one each time this is passed. Fifty turns are required. Leave about 6 in. of wire at the start and finish of the coil, to make the necessary connections to the binding posts. When the coil is completed, scrape off the insulation from the wires on one of the spokes for a space about ¼ in. wide, running straight down the spoke toward the center. Put the spiderweb form on the center of the baseboard with the scraped section of the wire pointing to the aerial post. Then place the crystal holder in the exact center of the spiderweb coil, with one of the screw holes for mounting the crystal holder in direct line with the scraped wire and the aerial post. Fasten down the holder to the base with a wood screw through the mounting hole on the opposite side of the crystal holder.

At the starting, or inside, end of the coil, drill a small hole straight down through the base, and on the under side, cut a groove in a straight line from this hole to the ground binding post. Run the inside end of the wire down through this hole and along the groove to the ground post. Also drill a small hole at the outside end of the coil, and another just in front of one of the binding posts on the crystal holder. Join these two holes with a groove, and bring the outer end of the coil down through the base, along the groove and up to the post on the crystal holder. Drill another small hole just in front of the other binding post on the crystal holder, and run a wire down from this side of the crystal, through a groove alo ʒ the bottom to one of the phone posts. The other phone post is then connected with the bottom of the aerial binding post in the same manner.

With a hacksaw, cut off a section of the square slider bar, long enough to bridge the space between the base of the crystal holder and the aerial post. Measure the distance from the screw hole in the flange of the crystal-holder base to the aerial post, and drill holes in the slider bar accordingly. A 6-32 brass machine screw, 1½ in. long, is brought up through the baseboard, and through the opening in the flange. This hole is drilled down through the flange, using the hole in the crystal holder as a template. The hole on the under side of the base is countersunk for the head of this screw, which is



The Completed Set; Dotted Lines Indicate Leads under Baseboard

MATERIAL LIST

1 wood base, ¾ by 6 by 6 in.
4 binding posts.
1 spiderweb form.
1 small spool No. 24 d.c.c. magnet wire.
1 slider bar, 3 in. long, with slider.
2 6-32 brass machine screws, 1½ in. long.
1 wood screw, 1 in. long.
1 crystal and holder, upright type.
Washers and extra nuts.

not shown in the illustration as no circuit wires go to it. A nut is fitted on the end of the screw protruding through the holder flange above, and the slider bar slipped down on the nut, where it is held by another nut. At the aerial post, a brass machine screw of the same size is passed up through the baseboard, the lead from the phone post first being securely fastened to the head of the screw on the under side. Washers or a bushing should be used to level the bar. Put the slider on the bar and mount on the screw. The aerial binding post is then screwed down, and the set is ready for use.

To operate the receiver, use a one-wire aerial, about 75 or 80 ft. long, and connect a lead from the ground post to a clamp attached to a cold-water pipe. Then insert the phone-cord terminals in their posts. Tune by moving the slider along the bar until a station is heard at maximum volume and adjust the crystal to the spot where best results are obtained.

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126

THE BOY'S BOOK of CRYSTAL SETS

BY

W. J. MAY

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PREFACE

The Publishers make no apology for reviving interest in the crystal set. For too long the technical press has immersed itself in an orgy of complex superheterodyne receivers and television equipment, all of which is necessarily expensive. In the 1920's, crystal set construction was a fascinating and inexpensive pastime indulged in by the majority of boys of that day. Today this pleasurable hobby can be even more attractive, the introduction of the germanium crystal diode and high performance coils, has opened up new paths which will capture the interest of boys of all ages. Perhaps most important of all. not only can a receiver capable of impeccable reproduction be produced without any technical knowledge; but the cost is no greater than that of a few visits to the cinema

INTRODUCTION

Before any attempt is made to construct a receiver, it is necessary to examine the problems which surround the crystal set, so that the best can be obtained from any of the designs attempted.

Firstly, it must be understood that the crystal set as it is to-day does **not** provide any amplification. It relies entirely on what is fed into it via the aerial and earth system and gives a very faithful replica of the original transmission.

From this it will be obvious that the aerial and earth system must be as efficient as possible if the final results are to be in any way outstanding. This because these are the only means by which the signals are fed to the receiver.

Secondly it is necessary to understand the nature of the transmitted signal, then it will be easy to understand the working of the set and to appreciate the function of each of the components.

When crystal sets first became popular, very little information was generally available, at least, not in a form that school-boys could understand.

We all built sets of all shapes and sizes, with coil designs that had to be seen to be believed, but very few of us had much idea of how they worked.

I well remember the case of a cousin of mine who, in those days acquired a magnificent variable capacitor or condenser as it was then called, this instrument had a most impressively engraved dial of polished ebonite, brass vanes and nickel plated end plates. Having observed similar dials on several highly priced commercial receivers in the town, he at once ripped out the somewhat tattered coil from the family receiver and replaced it with this device. The profound silence which ensued caused considerable amazement and dismay until a better informed adult explained the mysteries of L and C to him. Readers of this manual, however, will be better informed and there is no risk of failure with any of the designs described provided the instructions are carefully followed.

The Signal

To commence, we will assume that an orchestra is playing in a broadcasting studio. Since the principle of radio transmission is electrical it is necessary to change the sound produced by the orchestra into an electrical equivalent. This is carried out by the microphone, which picks up the sound and changes it into minute electric currents.

As they appear at the output of the microphone they are too small to be of use and accordingly are passed through a high power amplifier. These amplified currents could now be transmitted, but unfortunately, as we shall learn, owing to the inherent nature of the signal in this state, transmission over any useful distance would be impractical.

When the music from the orchestra is transformed into electrical currents they are in the form of alternating currents, usually called A.C., that is they rise to a maximum in one direction, fall to a minimum, rise to a maximum in the opposite direction and then fall to minimum again. This process is repeated over and over again. One complete rise and fall in each direction is called a cycle and is drawn in Fig. 1. Every time a note is struck on a piano, vibrations are sent out which reach the ear enabling you to hear it. These vibrations are also spoken of as cycles, they rise and fall in intensity the same way as an alternating current. The number of cycles radiated by any given note over a period of one second are referred to as its frequency. Middle C on the piano sends out 261 cycles every second and is known as having a frequency of 261. The microphone also "hears" the note and in the case of middle C produces minute A.C. at 261 cycles. This can be drawn as in Fig. 2, the only difference between Fig. 1 and 2 is, that the time factor is given so that the frequency can be identified. The higher the pitch of a note the higher the frequency and the lower the pitch the lower the frequency. On a piano the frequency of the top note is 3515 cycles and that of the lower 27 cycles. Those of you who have listened to an organ in a concert hall will have noticed that when a very deep note was played, it sounded like a growl to the ear, but the vibrating fre-

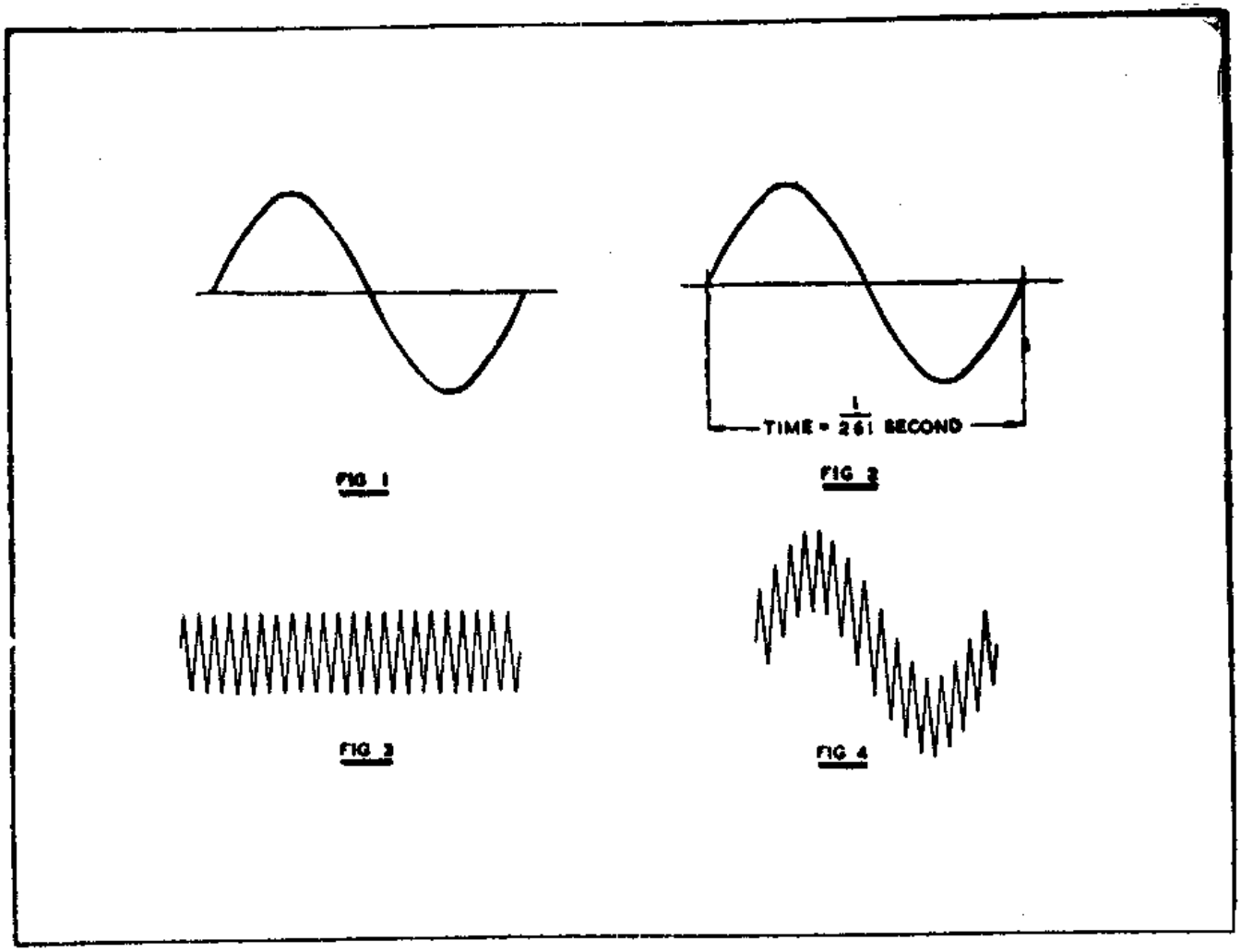


FIG 1

FIG 2

FIG 3

FIG 4

quency could be distinctly felt through the seat. Higher notes have too high a frequency to be observed in this manner.

The range of sounds which can be detected by the human ear are known as audio or low frequencies. From this you will understand that a low frequency amplifier is one which amplifies sound.

So that the transmitter will carry the programme over a useful distance it is necessary to radiate high frequencies. Now, as explained, the programme to be transmitted consists of low frequencies, and to overcome the difficulty, the transmitter generates A.C. of high frequency and combines it with the low frequencies. It will now be understood that the transmitted signal consists essentially of two different parts, a high and a low frequency content.

Fig. 3 gives a representation of the high frequency signal generated by the transmitter. In the case of the London Home Service, the frequency is 908000 cycles.

When referring to a high frequency signal on the medium or long wave-band it is usual to express the frequency in thousands of cycles, thus 908000 cycles becomes 908 kilo-cycles, which in turn may be abbreviated to 908 k/cs. It might at first be thought that by adding the low frequency or L.F. signal to the high frequency or H.F. carrier a form such as in Fig. 4 would result, such a combination is useless, and, so that the original L.F. content can be satisfactorily extracted by the receiver, the L.F. signal must vary the amplitude or output power of the H.F. signal as in Fig. 5.

It is in this form that the signal arrives at the receiving aerial. The aerial in itself is incapable of discriminating between one signal and another, and countless signals will be collected by the aerial at any one time. Many of these are too weak to be of use but the stronger ones must be sorted out since there is no point in receiving several programmes at once.

Fig. 6 shows the basic circuit of the input to a crystal. The coil L possesses a quality known as inductance, and the capacitor C, that of capacitance. If the coil had no capacitance whatever across it, all signals arriving at the aerial would be effectively short circuited to earth. As a matter of interest, it is impossible to obtain this state of affairs since even without any additional capa-

citor any coil **must** contain a certain amount of self capacitance.

By combining a coil and capacitor as in Fig. 6 a peculiar effect is observed, at one particular frequency, the signals are not short circuited to earth, but are developed across the coil. In other words the combined effect of L and C no longer provides a short circuit, **but only at one particular frequency.** If the value of C is altered the effect will be observed at a different frequency; likewise by altering L the frequency at which the effect will take place can be changed. There is a name for this phenomena, the frequency at which it occurs with any given L and C combination is known as the **resonant frequency.**

The values of the coils and capacitors shown in this manual have been carefully chosen so that resonance will be obtained at all frequencies where stations are broadcasting. Broadcasting stations work in bands of frequencies, those of major interest to crystal set constructors are the medium wave-band 1200 k/cs—600 k/cs and the long wave-band 300 kc/s—150 k/cs.

Usually a variable capacitor is used with a fixed inductance to cover one band and an additional coil switched in to increase the inductance to cover the other. In this way the L and C combination can be adjusted to provide resonance at the desired frequency of any given station. In other words you can **select** the station you want by varying C, that is turning the dial of the variable capacitor. This procedure is referred to as tuning. Having selected or tuned the required station it still remains necessary to change the form of the signal back to that of the original transmission. This process is called detection or de-modulation.

Examination of Fig. 5 will show that the signal has been duplicated, in other words, as it rises in one direction it also rises equally in the other. In this form the signal is useless since each half of the signal cancels the other, and if this signal is applied to a pair of headphones silence will result. Obviously some provision must be made to get rid of the unwanted half of the signal, and it is here that the crystal detector must be considered. This device will pass current in one direction only, ignoring any signal in the opposite direction, so that if the signal of Fig. 5 is passed through such a crystal, that of Fig. 7 will result. There is still the H.F. content to be reckoned

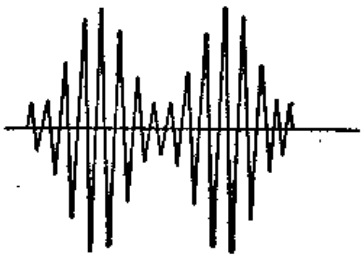


FIG 5

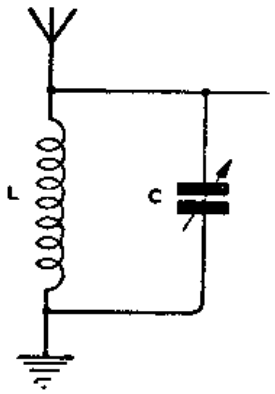


FIG 6

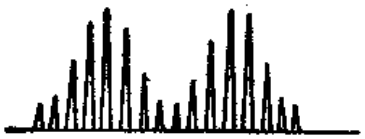


FIG 7

FIG 8

with, fortunately this is easily dealt with, a capacitor connected across the headphone terminals effectively disposes of this, leaving only the audio or L.F. content, as shown in Fig. 8. This audio content, which is a faithful replica of the original transmission is fed to the headphones. These in turn reverse the process of the microphone and transform the electrical currents into sound waves acceptable to the human ear.

Briefly then, your requirements are as follows:—

(1) A good aerial and earth installation, to make the most of the available signals.
(2) A receiver containing:—
  (a) Some form of coil and capacitor (L & C) combination to select or tune in the wanted station.
  (b) A crystal to get rid of the unwanted half of the signal (detection).
  (c) A fixed capacitor across the headphones to get rid of any remaining carrier.
(3) A pair of sensitive high resistance headphones.

Aerials

By this time the intending constructor will be able to appreciate the necessity of a good aerial. It is a point which cannot be over-emphasised. Assuming that you are in the fortunate position of being able to erect an outdoor aerial there are two main considerations, height and length.

One of the best that can be used is the inverted L shown diagramatically in Fig. 9.

It should be erected as high as is practical, every foot counts. The horizontal wire, that is the aerial proper, should have a minimum length of 60′ to which of course the length of the down lead is added. Where it is impossible to erect an aerial with an ideal horizontal length, a compromise must be effected.

Fig. 10 shows a three wire spreader aerial which gives quite a good effective length.

Suitable wire for a receiving aerial will not set any problems, stranded copper about 7/22 gauge is the best. 7/22 means that it consists of 7 strands of 22-gauge wire. This wire may be obtained covered, and for the present purpose is better than the plain or enamelled kind.

Note that insulators are used between the actual aerial wire and its anchoring supports, it is important that these are used, otherwise leakage will occur which will of course spoil its efficiency. Fig. 11 and 12 show how the wire may best be attached to two of the most common types of insulator available. Fig. 11 is of porcelain and is usually referred to as an egg insulator, whereas Fig. 12 shows a more modern (and more expensive) type in glass. If you are using the glass pattern, one is usually sufficient at each end of the aerial, but with the egg type two should be used. No doubt many readers will not be in a position to erect an out-door aerial, and must necessarily be content with an indoor installation.

The next best thing to a good out-door aerial is a replica constructed in a loft. If this form of construction is used, care must be taken when feeding the down lead to avoid close contact with the wall of the house. At the point where the lead feeds under the eaves, a length of rubber tubing can be used to cover the wire. The lead is fed through one of the small spaces left for ventilation purposes. A general idea is given by Fig. 13.

A less elaborate but quite effective aerial can be obtained by using a bed-spring. A length of 7/22 copper insulated wire is connected between the receiver and the spring. The spring should first be cleaned with emery cloth. Remove about 2″ of insulation from the wire and bind it tightly round the prepared spring. The joint may be covered with insulating tape which is obtainable from all electrical stores for a few pence.

Such aerials are quite popular since so many crystal sets are built for bedroom use.

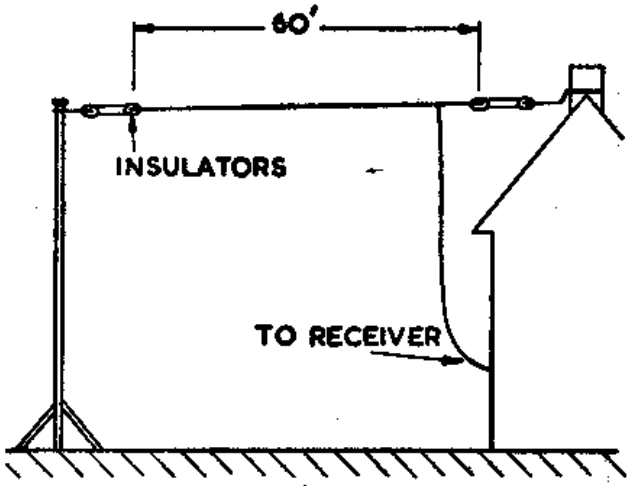
When an aerial is required in the living room, the picture rail can be conveniently used. Insulated screw-eyes are fixed at intervals of about 3 feet along the rail, the wire is firmly anchored to one of them, and stretched right round the room until you arrive back at the starting point. By use of one of the insulated screw-eyes, the wire is secured and the down lead fed to the receiver. The idea is illustrated in Fig. 14.

Soldering

Before leaving the subject of aerials, a few words on soldering will not come amiss. Down leads on outdoor aerials should be soldered, and the same applies to loft types. Apart from aerial leads; earth leads and all the connections in the receiver will require soldering.

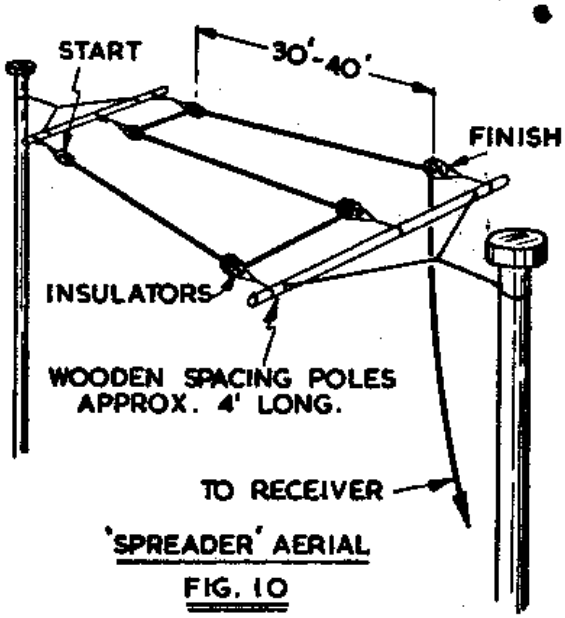
For the type of soldering necessary in radio construction an electric iron is the best solution. One of the small types marketed by Adcola or Henleys will be found admirable for the job. These are excellent for actual set construction but are hardly large enough for soldering the down lead to an outdoor type of aerial. Here the heat is dissipated much more quickly and a larger iron is required in order to get the solder to flow.

Ordinary irons which may be heated by a gas flame can be obtained very cheaply from most ironmongers' stores. Assuming the use of such an iron, first heat the iron until the copper bit is giving off a green coloured flame, the iron is now at the correct operating temperature. The bit is now discoloured or oxidised and its tip



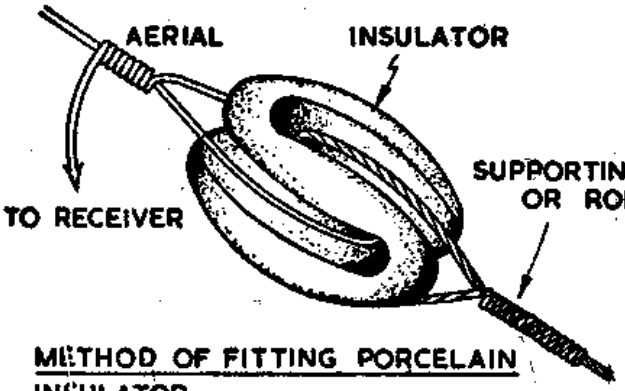
INVERTED 'L' AERIAL

FIG. 9



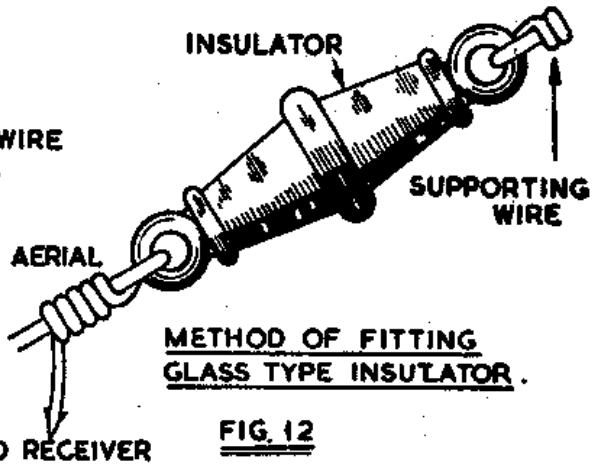
'SPREADER' AERIAL

FIG. 10



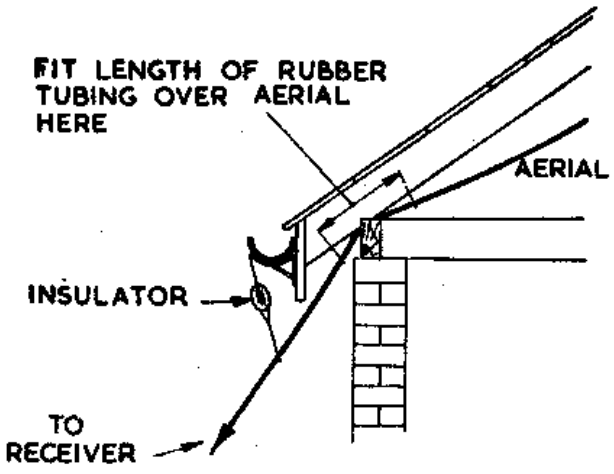
METHOD OF FITTING PORCELAIN INSULATOR

FIG. 11



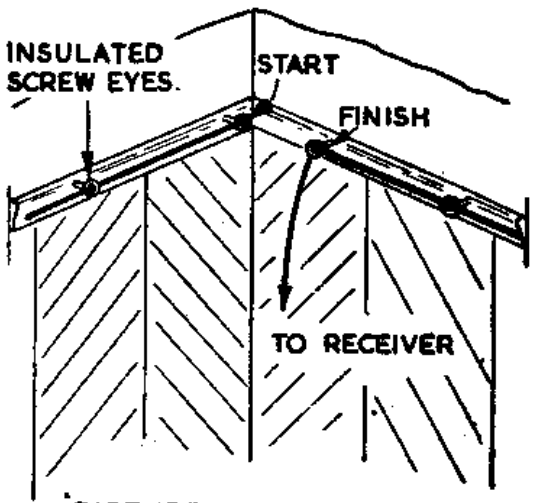
METHOD OF FITTING GLASS TYPE INSULATOR.

FIG. 12



LOFT AERIAL.

FIG. 13



'PICTURE RAIL AERIAL.

FIG. 14

should be quickly cleaned with an old file which should be kept specially for the purpose. A better idea is to obtain a small block of sal-ammoniac and rub the tip of the hot iron on it. This will clean the tip of the bit perfectly. Next take a length of cored solder such as Ersin Multicore, and melt a little on to the prepared tip faces, now smooth it evenly over the surface with a piece of old rag (be very careful not to burn your fingers) the iron is now "tinned" and ready for use. When heating the iron, be very careful not to let the bit overheat or get red hot, otherwise the tinned surface will be destroyed and the whole process will have to be gone through again.

An electric iron will not overheat, and the tinning will last much longer than with ordinary types, and since the heat does not deteriorate there is less likelihood of making faulty joints.

Having obtained a tinned iron, the process of soldering joints is remarkably simple and anyone with a little patience can acquire the art in a very short time.

To solder two copper wires together: clean the wire with emery cloth, apply a prepared iron and some cored solder to it, the solder will flow evenly over the wire thus tinning it. Repeat the process with the remaining wire and then twist the two together. Now apply the iron to the joint from the underside and the cored solder to the joint on top. Solder will flow evenly over the joint. Remove both iron and solder and allow to set. The solder will harden or set in a few seconds but during this period the joint must not be touched or moved, as otherwise the joint will be "dry" and quite useless mechanically or electrically. On radio components, tags for soldering are already tinned though if they are old, or discoloured it is best to re-tin them. Normally, however, it is only necessary to twist the connecting wire to it and apply the iron and solder as explained.

Remember: **never** apply the solder to the iron and then the iron to the joint, always apply the iron **and** the solder to the joint. It is however a good thing to apply a little solder to the iron tip even when it is perfectly tinned just before making a new joint.

One final "DON'T," you will have noticed that the solder referred to is "cored," that is, it contains resin and other substances through its centre, plain solder as used by electricians and plumbers will not do, as it is the resin or flux

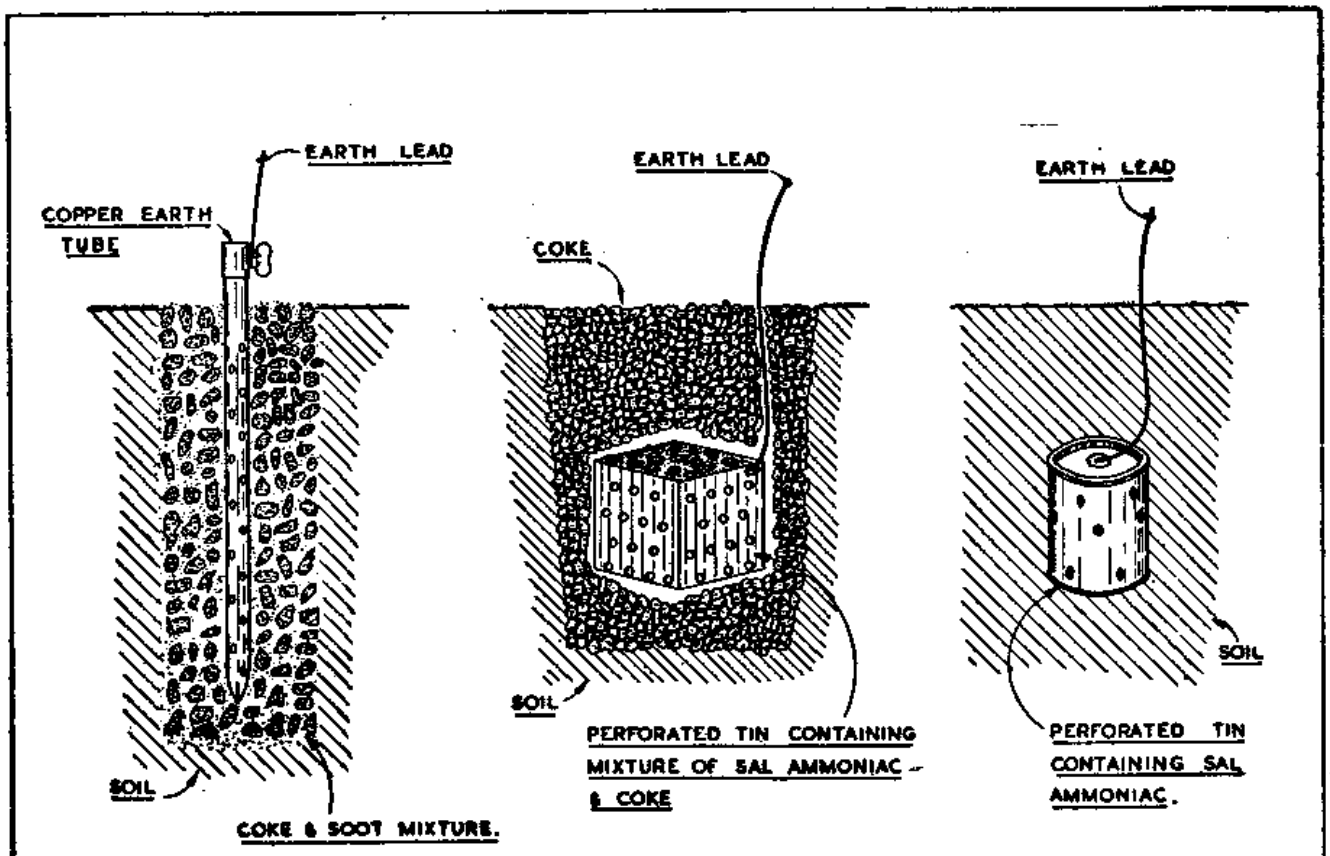


FIG. 15.

as it is called which makes the solder flow evenly and permits a good electrical joint.

Equally important do not attempt to use a separate flux or soldering fluid with plain solder or to "help" the cored solder, since these are almost certain to cause eventual corrosion and will destroy any components that have been contaminated.

Earths

The provision of a good earth is just as important as the rest of the installation.

Much disappointment would be avoided if this fact were not lost sight of, to avoid any slip up in this direction I am proposing to outline several well tried and efficient earth systems.

If it can be obtained, an earth rod, specially designed for the purpose provides the basis of a good earth connection. These were very popular some years ago when the majority of receivers were either crystal or battery operated, but in these days of modern mains driven sets they are not often used and consequently not always readily available.

Fig. 15 shows how it is used. First excavate a hole to a depth of three feet and fill with a mixture of soot and coke. Drive in the earth rod, which is a hollow copper tube, perforated. The lead to the receiver is connected at the top, and should be of covered 7/22 gauge, as used for the aerial system. Though not obvious it is important that this lead-in should be in covered wire, otherwise a number of indifferent earth contacts are likely to be made at various points along its length until it reaches the set. This is very undesirable and will spoil the efficiency of the system. It is essential to keep the soil surrounding the rod moist, which is one reason why the tube is hollow, and care must be taken to pour water into the tube at intervals.

A very efficient earth is the percolative type also popular at one time. Due to the chemicals used, it will extract moisture from its surroundings, thus maintaining a permanently moist earth.

It should be installed as in Fig. 15b.

The container is of copper or zinc, anything else will quickly rust away, again a good quantity of coke is used. Fill the container with a mixture of sal-ammoniac and coke, and then bury in coke as illustrated. This earth will not require further attention. It is possible to use powdered calcium chloride instead of sal-ammoniac but unless a chemist can be persuaded to make some up it is better to stick to sal-ammoniac.

If a zinc container is not available and you do not use or cannot easily get coke, Fig. 15c should be used. This makes a better earth than many so-called "earths" that I have come across. Obtain as large a tin as possible, make a number of holes as shown. Solder the lead in the bottom and fill it with sal-ammoniac. Replace the lid and bury in the ground. The tin will eventually rust away, but the replacement cost is negligible.

If it is quite impossible to make direct contact with the ground, a water pipe must be pressed into service. This should be a main pipe feeding straight to ground and not a hot water pipe or one fed from a tank. Scrape the pipe clean and twist the lead in tightly around it, a copper clip is even better. Do not attempt to solder on to the water pipe; since cold water is flowing through the pipe, it is extremely unlikely that your soldering iron will heat up the water supply sufficiently to allow a sound electrical joint, though you may spring a leak.

No attempt should be made to utilise gas pipes; the possibility of causing a fire is certainly very remote but, they make incredibly bad earth connections due to a number of joints made before true ground is reached. These joints are at best only semi-conductors, at least from an electrical standpoint.

Headphones

Since the late war there have been a large number of head-phone sets available on the surplus market. These may be roughly divided into two types, high-impedance and low-impedance. For the crystal sets detailed in this book high-impedance 'phones are required and the low-impedance pattern will not be suitable unless a matching transformer is used. As this is likely to cost more than the rest of the installation including the set, it will be as well to avoid them.

High-impedance types have an impedance of 2000Ω to 4000Ω whereas the low-impedance types are usually 600Ω.

The remaining consideration is weight; often cheap headphones are very heavy and uncomfortable to wear, every endeavour should be made to obtain 'phones as light in weight as possible.

Crystals

All the sets shown in this book have been designed to work with modern germanium crystals rather than the older galena crystal.

These germanium crystals require no adjustment, which, in itself, removes the main objection to this class of receiver. Suitable crystals are available from the following manufacturers and on the surplus market: Mullard, G.E.C., Brimar, Westinghouse and B.T.H. They are of robust construction, some in glass, others in ceramic or plastic, but must not be subjected to heavy knocks, otherwise the contact point may become dislodged.

From the information given, you should now be able to install an excellent aerial and earth system, and have some idea of how the crystal set works, so it is time to pass on to the actual receiver construction.

CONSTRUCTION 1

If you examine the following circuits you will find that each one is different. In most cases the difference lies in the coil design and/or the method by which the crystal and aerial is tapped into it. Each of these circuits has its own particular advantage to suit different conditions and the ideal circuit in some localities is not necessarily the best in others. It is not just a matter of a given circuit giving louder results than another, if it were there would be no point in showing more than one.

The main problem is to obtain adequate selectivity without reducing the volume level.

A receiver is said to be selective when it tunes sharply, a set with poor selectivity allows the stations to spread over the dial and when used near a transmitter will receive the local stations mixed together, which of course is useless.

Consider Fig. 16a, this is a very simple receiver, with no special attempt to provide any great amount of selectivity. In areas where signal

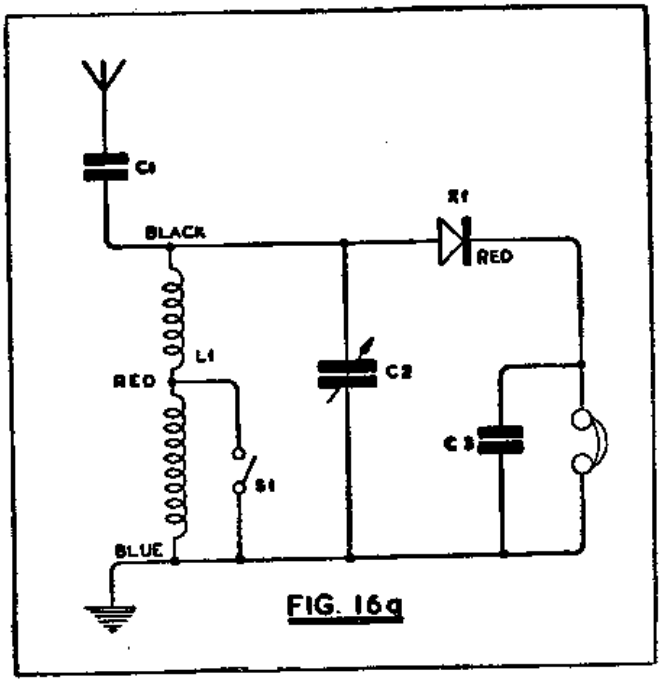
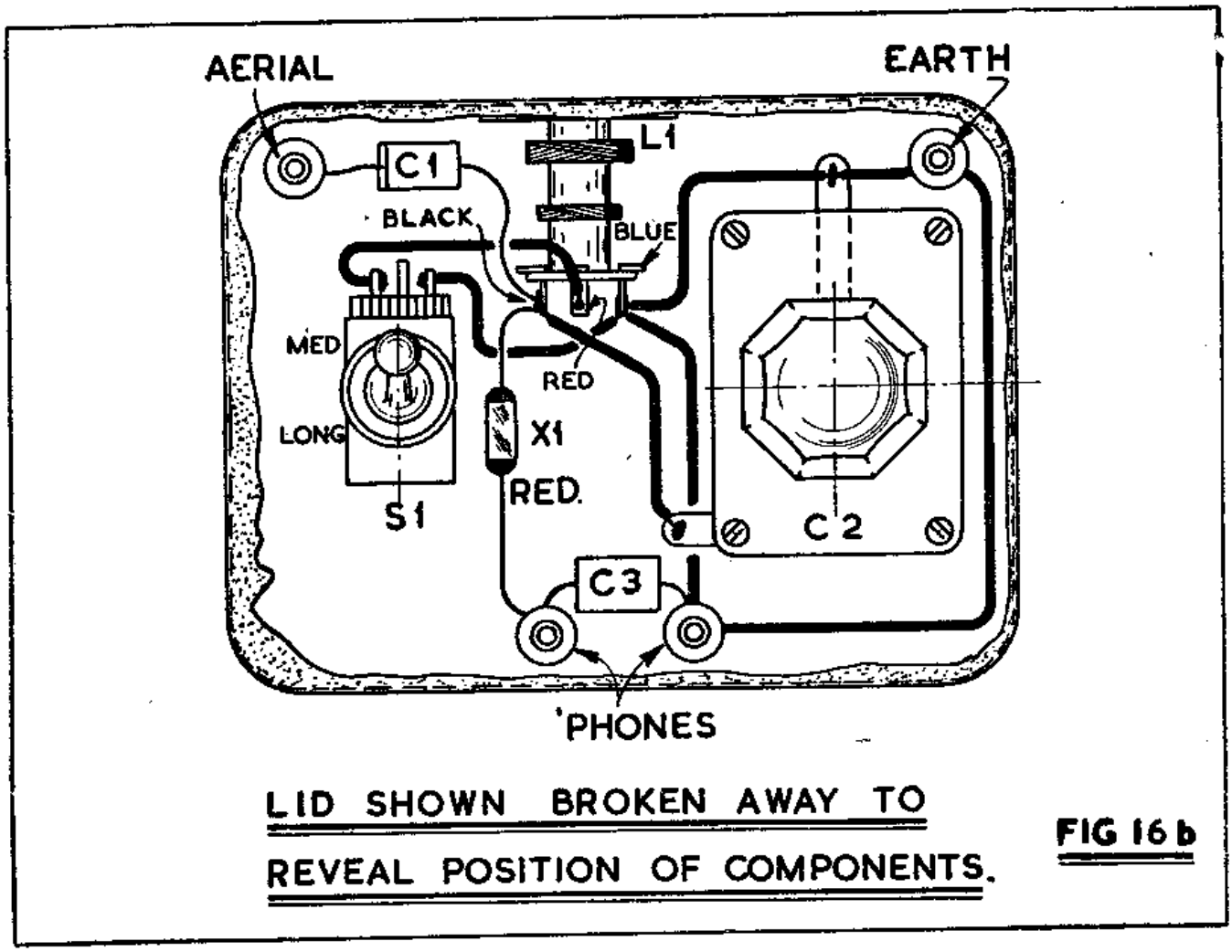


FIG. 16a



LID SHOWN BROKEN AWAY TO REVEAL POSITION OF COMPONENTS.

FIG 16b

strength is not high, or a short aerial is used, it will probably be ideal.

It would have been quite easy to increase the selectivity by providing a tap on the coil for the aerial as in Fig. 17a, but unfortunately, as the selectivity is increased overall volume is likely to decrease, so that unless you live close enough to the transmitter to have a large signal available and therefore need the selectivity, the circuit of Fig. 16a will be quite satisfactory.

Capacitor C1 is to prevent the aerial damping the circuit too heavily because this would flatten the tuning unnecessarily, however with small aerials it may be better to take the aerial direct to the black tag on L1. The signals are selected or tuned by L1 and C2, X1 is the crystal and C3 the capacitor across the 'phones to prevent unwanted carrier or R.F. reaching the 'phones.

A practical diagram (Fig. 16b) is provided showing the layout and all the wiring. You will require nuts and bolts to fix the coil, about ½", 4BA size will do, the other parts have locking nuts provided. An old 2-oz. tobacco tin makes a very good container and keeps the size down. Note that the metal box is connected to earth. With the exception of AC/DC receivers the metal work on any receiver or amplifier is connected to earth.

When S1 is open as in the diagram, the set will tune in long-wave stations, but when closed the medium wave-band will be received.

Components List, Fig. 16a

C1	100pF mica capacitor.
C2	500pF tuning capacitor, solid dielectric.
C3	1000pF mica capacitor.
X1	Germanium crystal.
L1	Crystal Set Coil. R.E.P.
S1	Single Pole toggle switch.
4	Insulated wander-plug sockets and plugs.
1	2-oz. tobacco tin (or similar container).

Make sure that the wander-plug sockets are of the **insulated** type, otherwise the metal case will join all the sockets together electrically.

Try to follow the theoretical diagram when wiring, a little practice will soon enable you to wire up a set without a practical diagram, which is a great advantage because often only the theoretical diagram is given when circuits are detailed in the technical press.

Fig 28 on page 34 gives a list of symbols used on the theoretical diagrams, so that you can readily identify the components.

2

Fig. 17a circuit is similar in many respects to that of Fig. 16a. The difference is purely one of selectivity, tuning will certainly be sharper, and even with comparatively inefficient aerials the design will put up a very good performance. There is no reason why the construction should not follow the same lines as the previous receiver but, by enlarging the set a little and using an air spaced tuning capacitor, the efficiency is improved. Observe that the colour coding on the coil, an R.E.P. Dual Range (Blue Box) is different from that of the Crystal set coil used on the previous design, and make sure it is correctly wired in.

C1 is a mica compression capacitor, often referred to as a "trimmer" and because it is adjustable, permits the set to be matched to aerials of varying lengths. Construction is carried out on a square panel of bakelite, perspex or wood, as shown in Fig. 17b. Perspex being clear like glass presents a most attractive finish, provided the set is neatly wired. It must be drilled slowly, however, otherwise the generated heat of the drill will make the hole wander, and spoil the panel.

After the set has been wired and tested a small wooden container can be made to house the completed set.

When carrying out reception tests, a little experiment is well worth while, to get the best out of the set. Try removing the germanium crystal

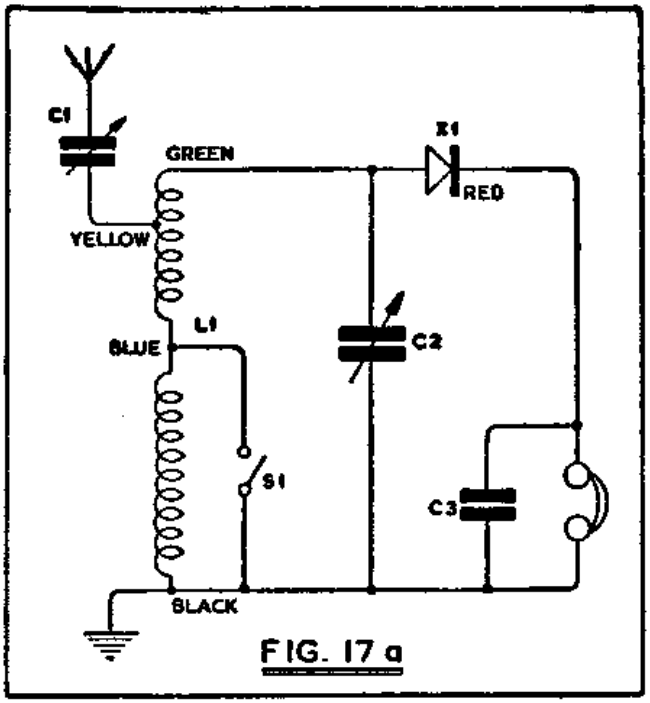


FIG. 17a

from the green tag on L1 and connecting it to yellow, at the same time removing the lead from C1 to yellow and connecting it to green. Once the best arrangement has been found the wiring

can be left permanently in that position. Note that two tags on the coil are unused, this is intentional because on this circuit the extra winding connected to these tags is not required. With the switch S1 closed, the coil will cover the medium wave-band, and when open, long wave stations can be received. It is customary to abbreviate the expression medium wave-band to M.W. and long wave-band to L.W.

Components List, Fig. 17a

C1	100pF mica trimmer capacitor.
C2	500pF variable capacitor (air spaced).
C3	1000pF mica capacitor.
L1	Dual Range Coil R.E.P. (Blue Box).
S1	Single Pole toggle switch.
X1	Germanium Crystal.
4	Terminals or wander-plugs and sockets. Perspex, bakelite or wood for mounting panel.

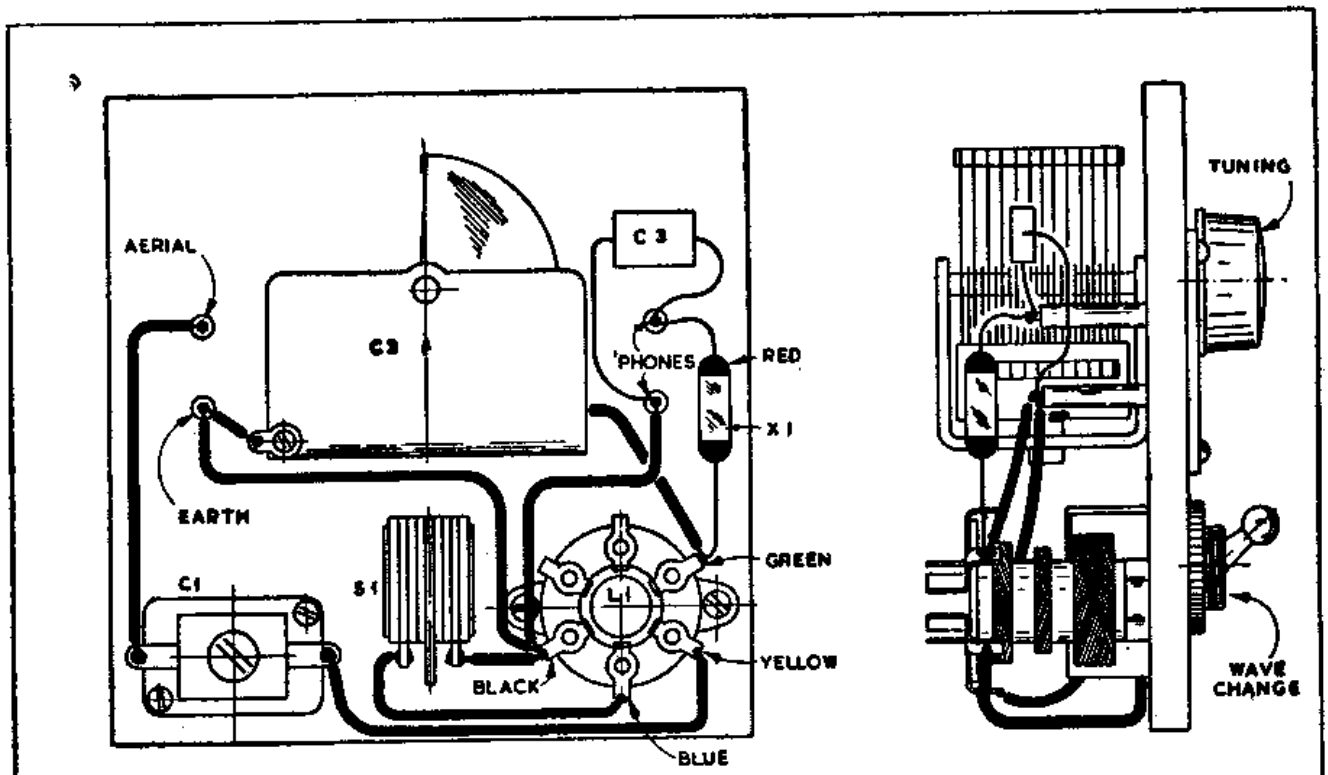
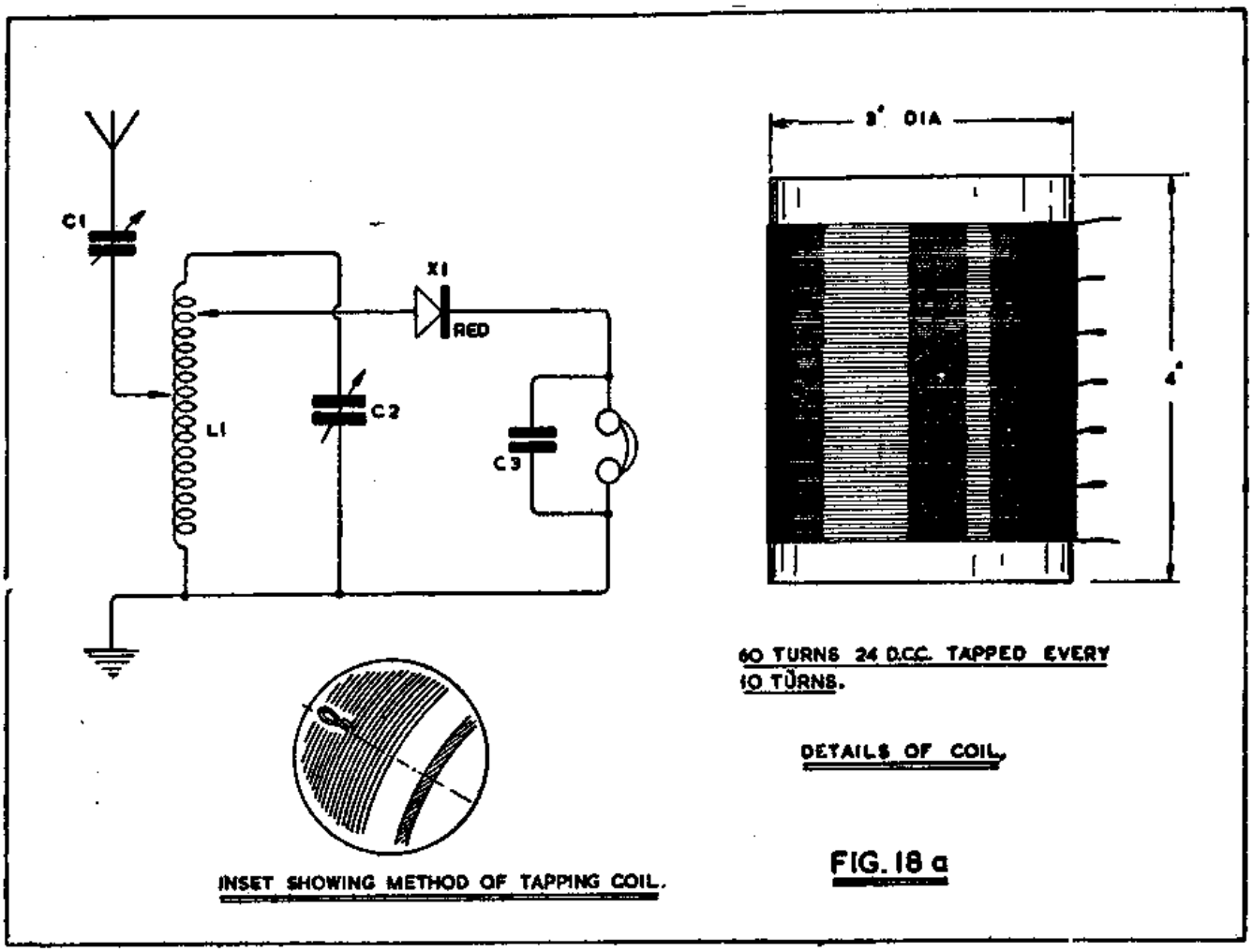


FIG. 17 b

3

The design shown in Fig. 18a and 18b is more flexible than either of the preceding sets. That is, it can be varied to suit local conditions to a greater extent than the first two sets. This is made possible by the large number of taps provided on the coil.

You will notice that in this instance the coil is home-constructed on a cardboard former and is much larger than commercially produced coils. Coils wound in this manner are called **solenoids**, the turns are wound on side by side; in this case to a depth of some three inches. The diameter of the coil is made purposely large so as to obtain high efficiency. Modern commercial coils are invariably wave-wound and quite often are litzendraht which is the German for litz wire. This wire is made up of a number of strands of fine copper wire, each strand is enamelled to insulate it from the others, the whole is then silk covered. Litzendraht is more efficient than solid copper wire, and by using this and wave-winding, manufacturers can produce an efficient coil which is also small. Unfortunately wave-wound coils cannot be produced without a complex winding machine and home-constructed coils must take the solenoid form. It would be possible to use litzendraht but it is not easy to obtain by the reel, it is most expensive, and is difficult to handle. This last point is because at termination points, each strand must be cleaned of its enamel before a joint is made, and if one strand is broken, its advantage over plain copper wire is lost. How-



ever, by using a former of reasonably large dimensions an efficient coil can be produced cheaply using ordinary copper wire.

The wire must be firmly secured at the start and finish of the windings. Pierce three small holes about ¼" apart ½" from the end of the former. Pass the wire through the first from the outside, return it through the second and pass it back again through the third. Leave some 6" of wire at the end to make off the connection. It will now be possible to wind the turns on tightly without wire slipping.

Count on ten turns and make a loop 1" long. The method of preparing loops is shown in Fig. 18a, loops or taps should always be made in this manner, never by baring the wire and soldering a further length of wire to it. Carry on with the winding, making off the taps every ten turns as directed until the coil is complete.

Construction is carried out on a wooden base-board and front panel. Secure the coil to the base-board as shown on Fig. 18b, mount the remaining components and proceed with the wiring.

When testing out, it will be found that the further the aerial is tapped down the coil towards the earth end, the greater the selectivity. A position should be found which permits separation of local stations without excessive loss of sensitivity.

The crystal tap is adjusted for best results and different settings of C1 tried out, when choosing the best position for the aerial tap.

Components List, Fig. 18a

C1	450pF padder.
C2	500pF variable capacitor.
C3	1000pF mica capacitor.
L1	See text and Fig. 18a.
X1	Germanium Crystal.
2	Crocodile Clips (to connect leads to coil taps).
4	Wander-plugs and sockets.

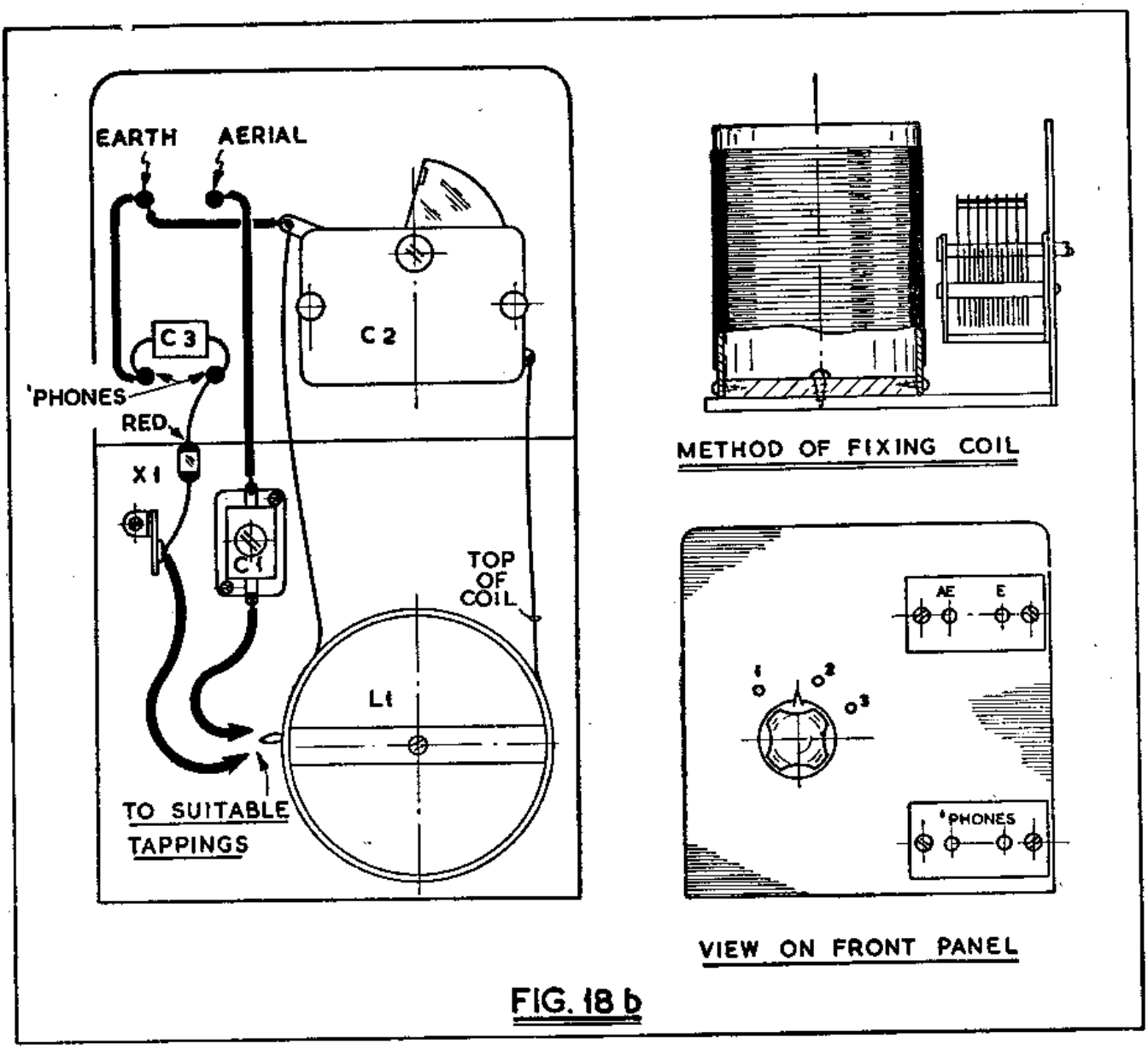


FIG. 18 b

4

A most unusual design is reproduced in Fig. 19a.

It first appeared in America about three years ago, and it certainly does offer some advantage over more conventional sets. The tuning circuits are duplicated and two crystals are used, so the circuit may be described as a full-wave receiver. Signals received will certainly be louder than with simple sets, but care must be taken with the coil winding, as with the other receivers using home-made coils, otherwise results will be disappointing. Be very careful to ensure that all three windings are in the same direction, this is very important.

The distance between each winding should be ¼". After the coil has been wound it is a good plan to warm it before a fire and paint the windings with "Durafix." Heating the coils makes the "Durafix" run freely. This substance sets quite hard and there will be no risk of the windings loosening. A small quantity of enamelled wire will cover the requirements of this coil, a 2 oz. reel will provide more than sufficient.

Before wiring in the germanium crystals, examine them carefully, note that one end is coloured red or in some cases marked with a positive sign thus +. You will notice, the sign is the same as the addition symbol used in arithmetic.

It is essential that both the red or positive ends are connected together, note that this is clearly marked on Fig. 19b. The receiver cannot work if

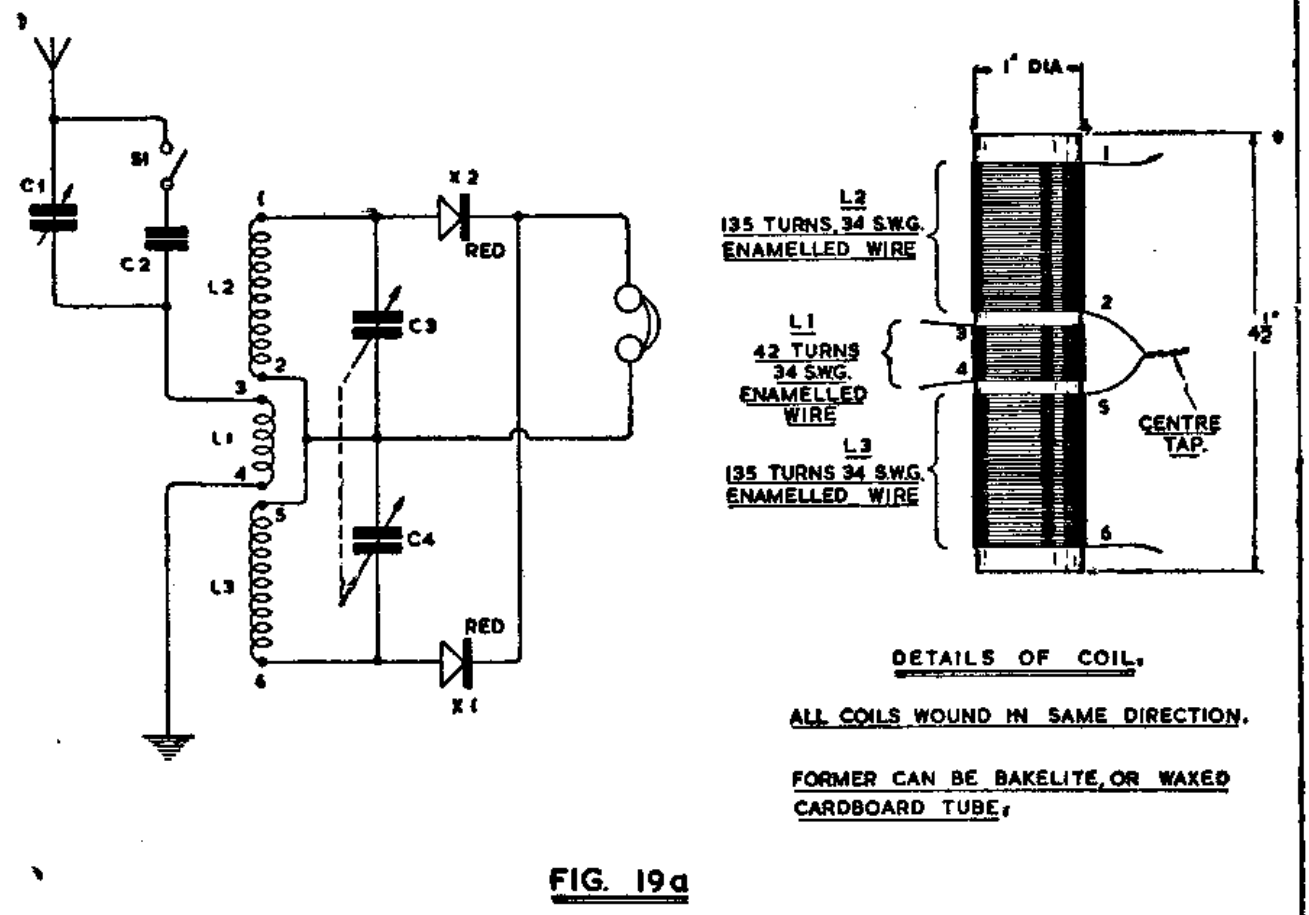


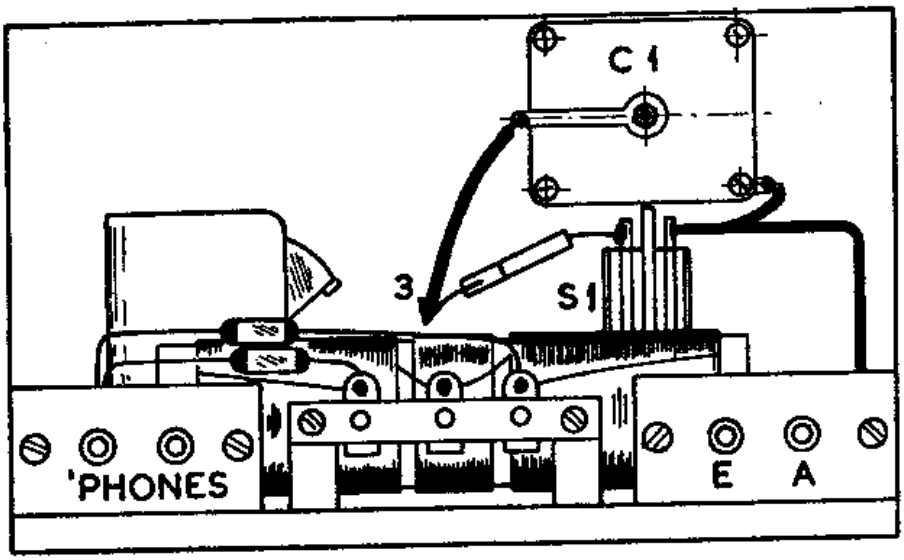
FIG. 19a

one of the crystals is connected the reverse way round.

A baseboard and panel form of assembly is used for this set which looks quite attractive if housed in a small polished cabinet. Dimensions are not at all critical but Fig. 19b gives a general idea of the layout that should be used. To avoid any unnecessary losses the coil should be mounted on perspex.

Obtain two strips of perspex 5½" long by ½" wide, drill fixing holes at both ends of each strip. It is best to clamp the two together when drilling these holes. Place one strip on the board in position, then put the coil and remaining strip over the first one and screw down. The coil will be clamped neatly and rigidly into position.

Twin socket bakelite strips are used to carry the aerial/earth and 'phone connections, if desired terminals mounted on pieces of bakelite could be used. These socket strips can be purchased from most shops selling components. If feet are not provided, drill two holes and screw into the edge of the base-board. Tag-strips are used to anchor some of the wires from the coil, and the germanium crystals. These also can be purchased for a few pence. At least two tags are necessary on one, and three on the other, though if the strips have more tags than required it is of no consequence. C3-C4, the ganged tuning capacitors have trimmers fitted, this should be stipulated when purchasing, and when first operating the set these small trimming capacitances are set about half-way. The idea of these trimmers is, that they compensate for any difference between the self-capacity of the tuning coils. With the switch S.1 open, that is in the OFF position, tune


C1
3
S1
'PHONES
E
A

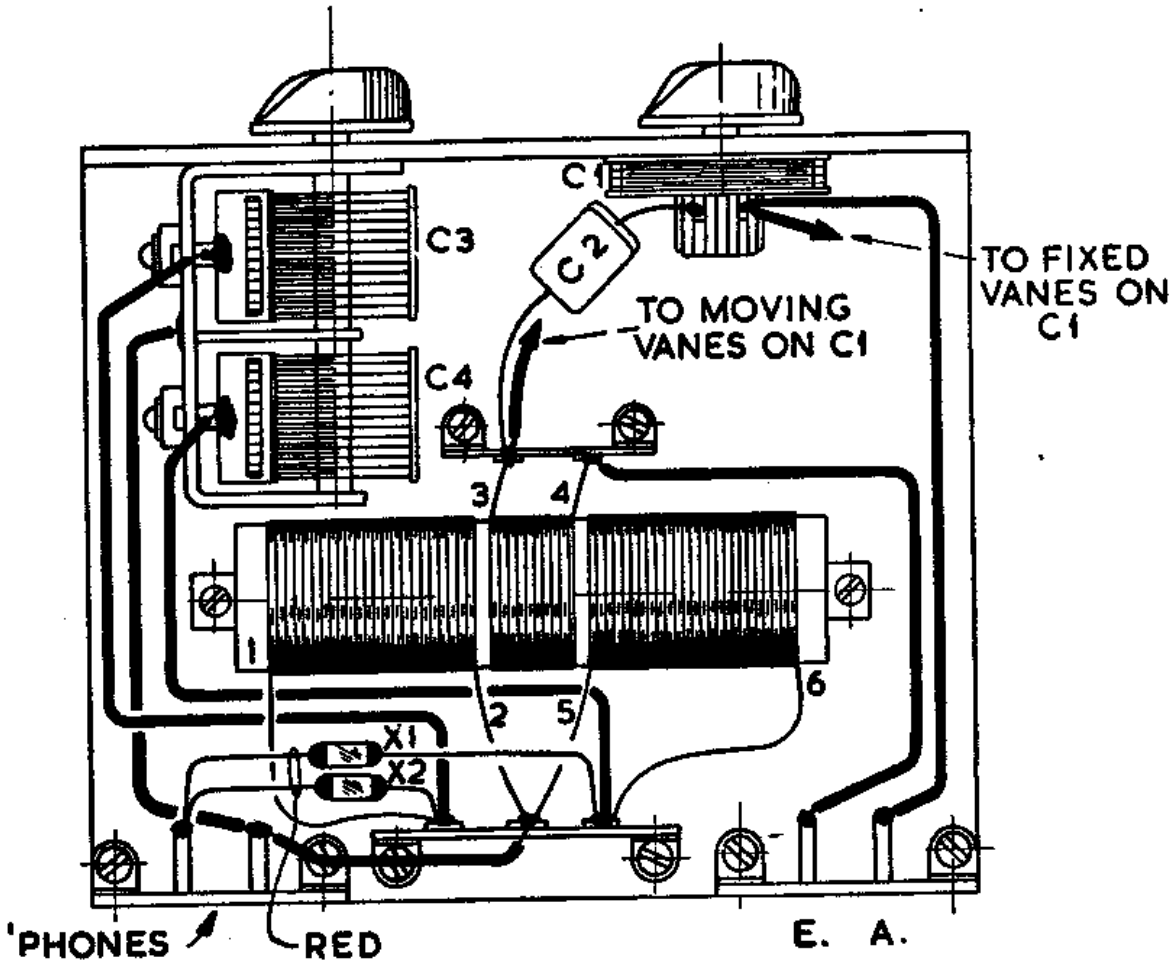

C1
C3
C2
TO FIXED VANES ON C1
TO MOVING VANES ON C1
C4
3
4
1
6
2
5
X1
X2
'PHONES
RED
E. A.

FIG. 19 b

in a station at the high-frequency end of the band, that is with the vanes of C3-C4 towards the disengaged position. Adjust C1 for maximum volume, without interference from other stations. Now adjust the trimmers for loudest headphone strength. It may be necessary to run over the adjustments several times for best results. Once they are correctly set, the trimmers require no further adjustment. The tuning coil covers the M.W. band, 1500 kc/s. to 600 kc/s., as the capacitor C3-C4 is advanced and the vanes start to mesh, the frequency to which the receiver tunes **decreases** so that with the vanes disengaged the set is tuned to 1500 kc/s. approximately and at full mesh, 600 kc/s. Switch S1 is used as follows, for stations **lower** in frequency than 850 kc/s it should be in the ON position with the contact closed, but for stations **higher** in frequency, it is in the OFF position. As an example both the London Home Service and London Light Programme transmitters on the M.W. band operate at a higher frequency than 850 kc/s. If any doubt exists as to the operating frequency of a given station, reference should be made to the Radio Times which quotes both wave-lengths and frequency. If you know the wave-length in metres, it is simple to find the frequency. Divide 300,000 by the wave-length in metres, the dividend equals the frequency in kilocycles, i.e., 300,000 ÷ 300 metres = 1,000 kc/s. In the same way, dividing 300,000 by the frequency in kilocycles will produce the wave-length in metres.

To use the receiver, tune in the signal by C3-C4, adjust C1 for maximum volume without allowing stations to overlap.

Components List, Fig. 19a

C1	300pF solid dielectric variable capacitor.
C2	1000pF mica capacitor.
C3-4	500pF twin gang variable capacitor.
L1-2-3	See text and Fig. 19a.
X1-2	Germanium Crystal.
2	Tag-strips.
2	Twin socket strips.
S1	Single Pole toggle switch.

5

Whenever crystal sets are discussed, constructors are apt to think in terms of medium-wave reception. When you consider that at least 98% of published circuits are designed for this band (sometimes with the long-wave band thrown in as an afterthought) it is understandable. However, if you can provide a good outdoor aerial (and an equally good earth) there is a lot of fun to be had listening to the short-wave bands. This receiver is designed specially for short-wave reception.

A metal chassis is used for construction, these can be obtained ready made in aluminium from most good supply houses. It can be quite small and on the original model a 6″ x 4″ was used. Low loss components are used so as to obtain greatest efficiency on the short wave bands. The coil is a commercial product which plugs into a 4-pin base. Best results were obtained on the 3mc/s to 7mc/s band though coils covering other bands are available.

Note that the tuning capacitor C1 is smaller than normally used and has a maximum capacitance of 140pF.

Coil type 706/R covers the 3mc/s to 7mc/s band, but the set is equally satisfactory on the M.W. band, and to cover this coil type 706P should be used. This coil has an iron dustcore which can be adjusted to make the coil cover the required band. The effect of the core is as if turns were being added or removed from an ordinary coil. When dealing with the short waves it is customary to refer to the frequencies in terms of mega-cycles (mc/s) rather than kilo-cycles. One mega-cycle is equivalent to 1000 kilo-cycles, that is 1,000,000 cycles.

Layout and wiring can be clearly followed from Fig. 20b and a theoretical diagram is shown in Fig. 20a. Care should be taken when arranging a mounting for the aerial terminal. The best plan is to cut a 1″ hole in the chassis, mount the aerial terminal on a piece of perspex and screw into the chassis. This avoids any unnecessary losses between aerial and chassis. The earth terminal is screwed directly into the chassis.

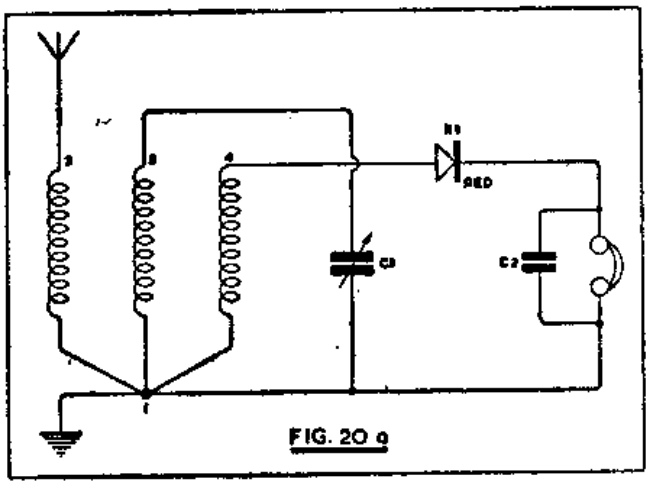


FIG. 20 a

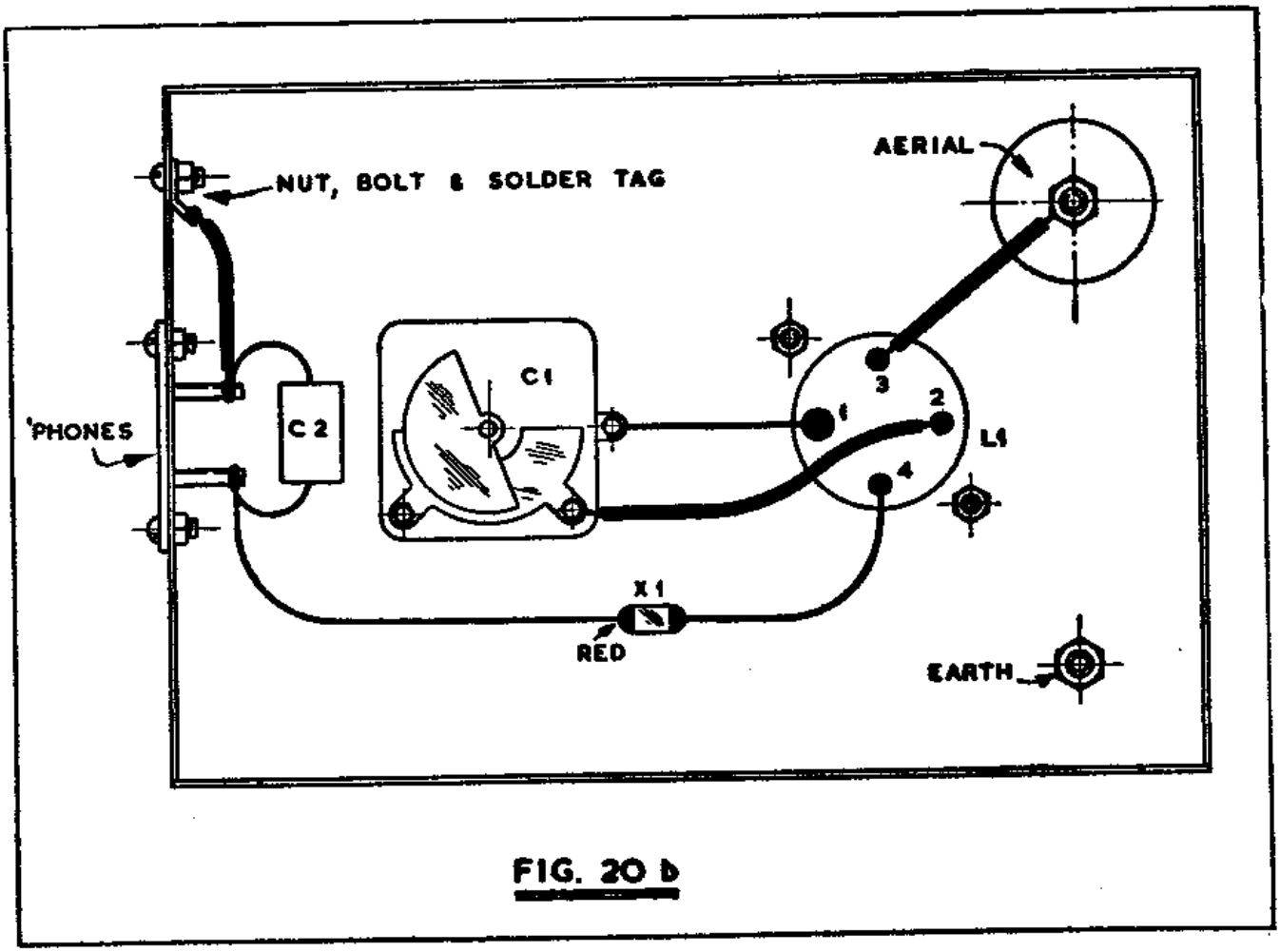


FIG. 20 b

Wiring as shown in Fig. 20a is likely to give the best results, but other arrangements are possible and to get the best out of the set they should be tried.

Refer to Fig. 20a. The aerial is connected to pin 3 on L1, C1 to pin 2 and X1 the crystal to pin 4. If you have a long aerial try this combination—Aerial to pin 4, C1 to pin 2, and the crystal to pin 3. Connections to pin 1 are not altered. In a few cases the following arrangement may prove best. Crystal **and** C1 to pin 2, aerial to pin 3, pin 4 left free and again pin 1 is unaltered.

Tuning on the short-wave bands is more critical than on the medium waves so tune very slowly over the band and remember signal strength is likely to vary from day to day.

Components List, Fig. 20a

C1	140pF variable capacitor Eddystone 586.
C2	1000pF mica capacitor.
1	Coil holder Eddystone 707.
L1	3m/c—7m/c coil Eddystone 706/R.
1	Engraved tuning dial.
1	6"x4"x2½" chassis.
2	Terminals (Aerial-Earth).
1	Twin socket strip ('phones).
X1	Germanium Crystal.

6

Fig. 21a-b is yet another design from America and appeared in Radio Craft some years ago, designed by Mr. W. J. Spain. The original used a silicon crystal which preceded the development of the germanium. Results are certainly very good though of course on the model built here, a germanium crystal was used. Selectivity can be adjusted to suit all conditions and even at short distances from local stations. Good volume can be obtained without the programmes overlapping. Home made coils are used and for best results a 75′ outdoor aerial is desirable.

Baseboard and panel construction is used, which can conveniently be housed in a small wooden cabinet.

Operation is a little more complicated than some of the other designs, but the results certainly merit the extra trouble taken. As with the other receivers much of the success obtainable is due to the coils and these must be carefully made otherwise results will be disappointing.

Layout and wiring can be followed quite easily from Fig. 21b and requires no special comment.

First make the coils. L1 is a tapped coil, the taps should be made in the same way as shown in Fig. 18b. It consists of 90 turns of 22 D.C.C. (double cotton covered) copper wire tapped at the following number of turns, 5, 10, 15, 25, 30, 40, 50, 60, 70 and 80. The former used has a diameter of 2". L2 is also wound on a 2" diameter former, this coil however is not tapped, it consists of 110 turns of 38 D.C.C. or enamelled copper wire. Great care must be taken when handling this wire, it is quite fine and in the hands of the inexperienced will easily break.

After wiring has been carried out the set can be put into operation.

Selectivity is controlled by S2, in position B tuning is broad but in position A it can be quite sharp.

When searching for a station switch S2 to B. Tune the station and switch S2 to A. The frequency range is controlled by the position of the tap on L1, a good plan is to start with the 5 turn tap on L1.

Components List, Fig. 21a

C1 500pF variable capacitor.
C2 500pF variable capacitor.
C3 1000pF Mica Capacitor.
X1 Germanium Crystal.
L1 & L2 See text.
S1 Single Pole toggle switch.
S2 Single Pole 2-way toggle switch.
2 Twin socket strips (Aerial/Earth and 'Phones).

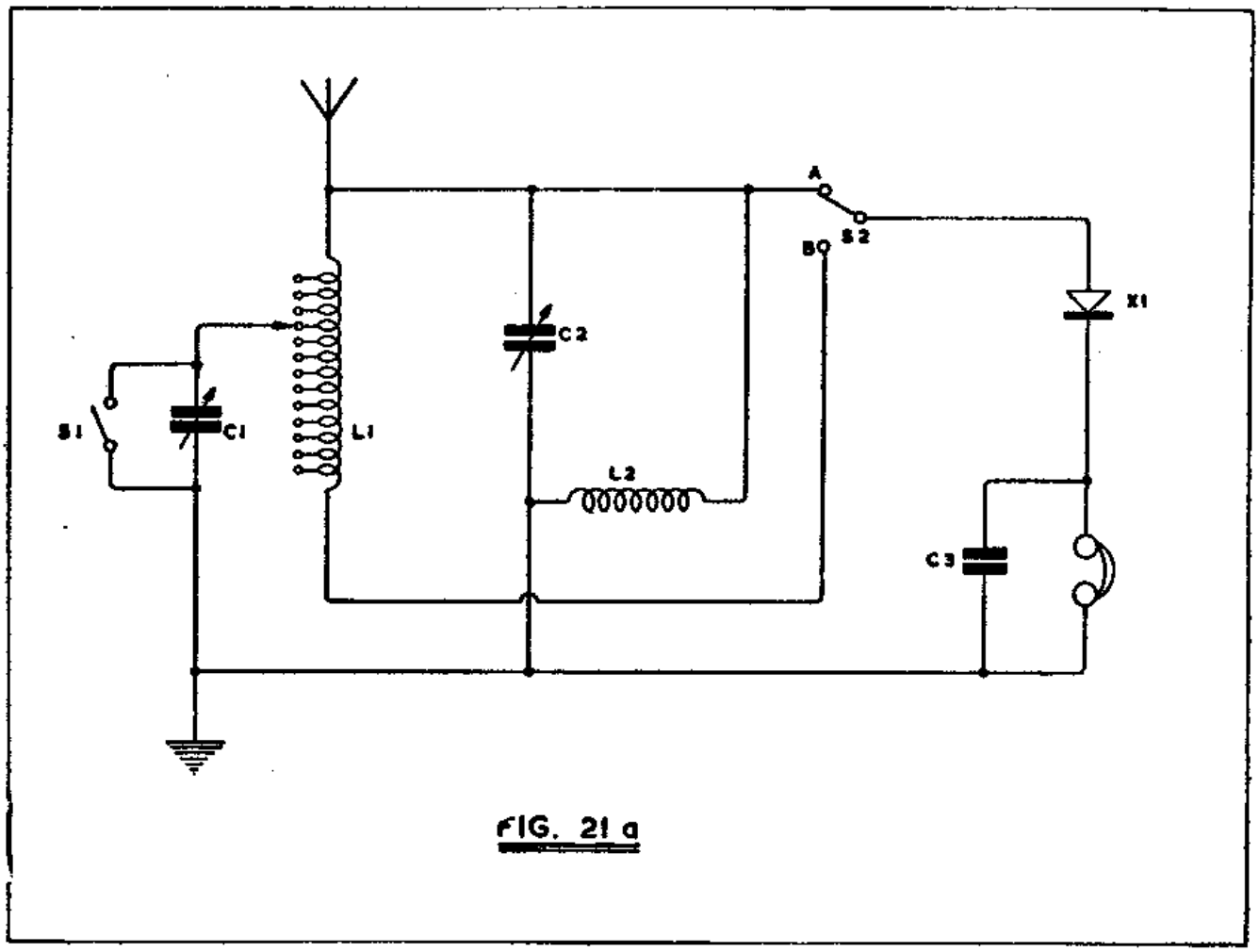


FIG. 21a

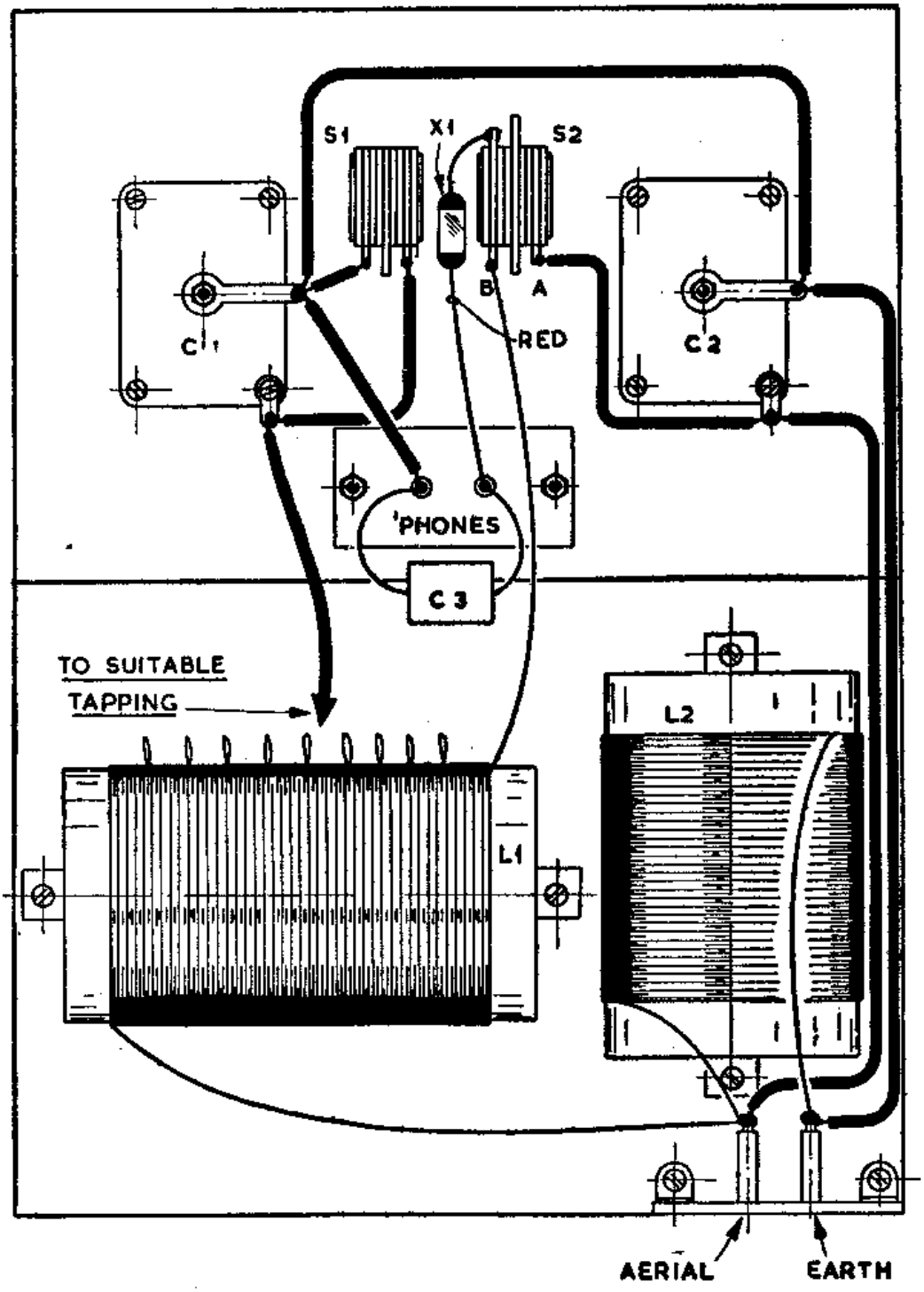

S1
X1
S2
B
A
RED
C1
C2
'PHONES
C3
TO SUITABLE
TAPPING
L1
L2
AERIAL
EARTH

FIG. 21 b

Keep S1 open and C1 for tuning, adjusting the tap on L1 for best results. Very powerful signals are best brought in with S1 closed. When using C1 and adjusting the taps on L1, switch S2 is kept in the A position.

It will be found that C2 acts as a fine control on C1. One division of the tuning dial on C1 is roughly equal to a movement of 20 divisions on C2.

Note that the coils are mounted at right angles to one another and that in this instance wood has been used as far as possible, an aluminium or other metal chassis is not suitable.

7

A very useful receiver for both medium and long-wave reception is possible with the circuit of Fig. 22a-b. A home-made coil is used. On the original a 1¼" diameter former was used but in fact this is not very critical and a 1½" former could be used. An interesting point is the construction of the long-wave coil section. Solenoid coils as used for the medium wave-band are not efficient if the length is too great relative to the diameter. Now as the long-wave section has 300 turns, if wound as a solenoid the winding length would be very great and a lot of efficiency would be lost. To overcome the difficulty the coil is pile wound. If the whole coil were to be wound in one pile it would still be inefficient because of the capacity formed in the winding, but by dividing it into five sections a coil of reasonable efficiency results. The same remarks concerning winding apply to this coil as to all the others in

Components List, Fig. 22a

C1 300pF Mica Capacitor.
C2 500pF Variable Capacitor.
C3 1000 Mica Capacitor.
X1 Germanium Crystal.
L1-2 See Text.
S1 Single Pole Toggle Switch.

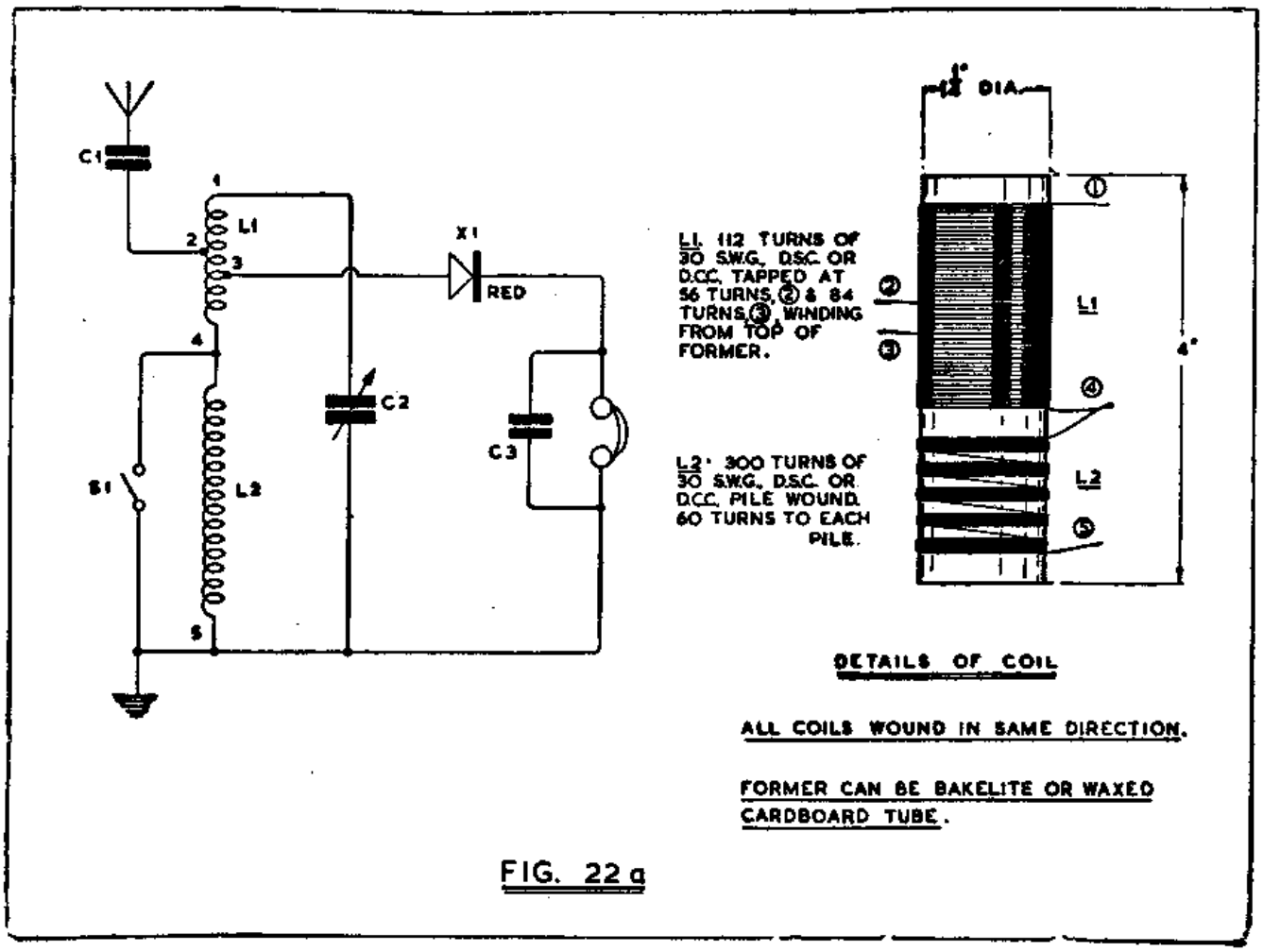


FIG. 22 a

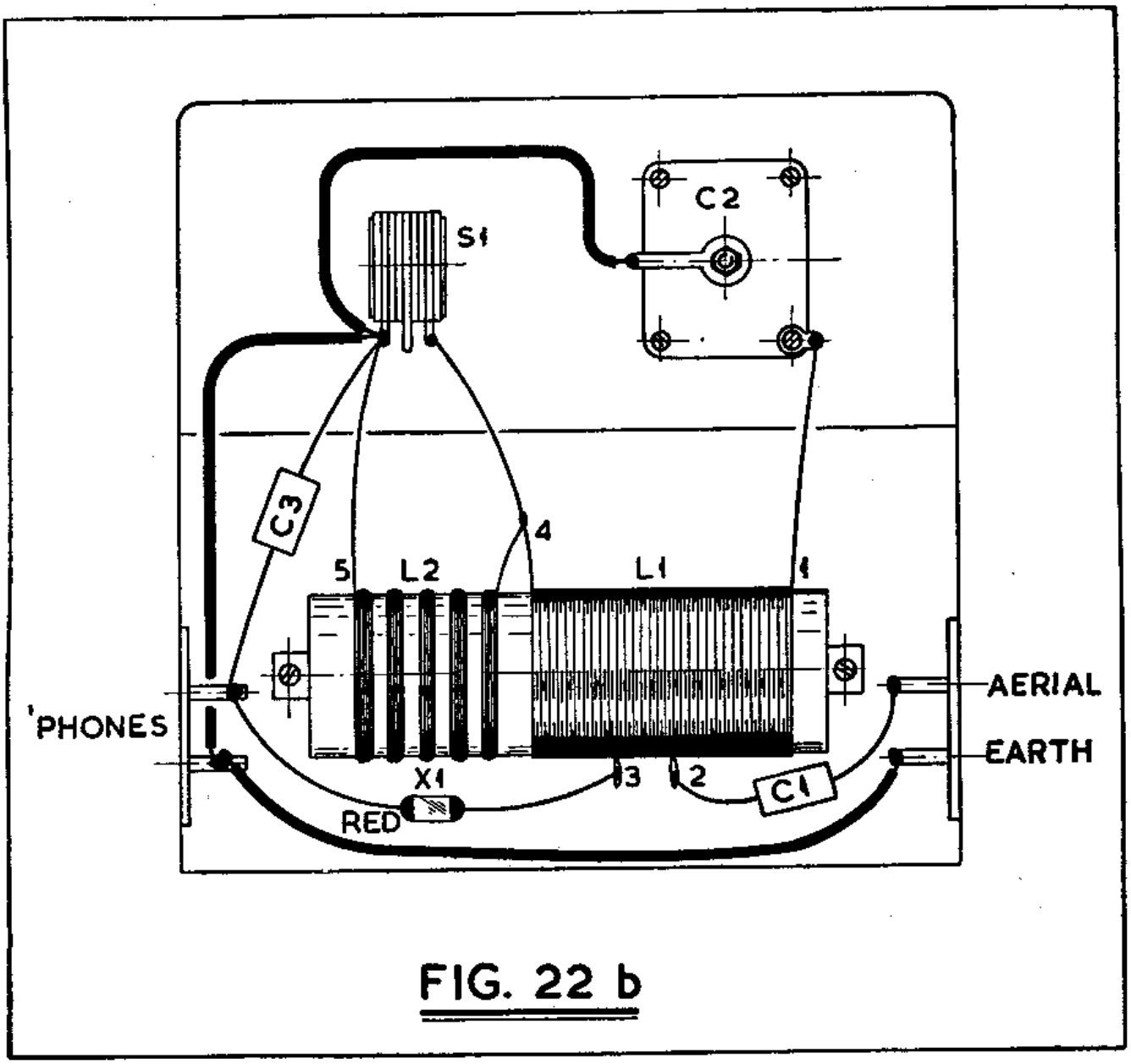


FIG. 22 b

this book, and readers should not, by this time, encounter any difficulties.

The design provides good selectivity with good sensitivity, and provided the set is not required to operate within a few miles of a powerful transmitter, no trouble will be encountered from interference between adjacent stations.

A wooden base-board and front panel is best for this design, and the wiring and layout can be clearly followed from Fig. 22b. Switch S1 is open for long-wave reception and closed for the medium-wave band.

Coil mounting can be on the same principle as for the full-wave design Fig. 19a-b.

8

If you look at all the other circuits shown in this book you will find that in each case the tuning capacitor and tuning coil are in parallel, this is by no means essential and the circuit of Fig. 23a shows a series arrangement that is the tuning capacitor is between the coil and earth instead of being wired up across it. This design also appeared in **Radio Craft** a few years ago.

An interesting point is that the coil is a conversion adapted from another type of component. To make this coil you must obtain an old I.F transformer as used in superheterodyne receivers.

You will require one designed to work anywhere between 450 and 470 kc/s. Some of the

very early ones worked at 110 kcs/ but these are unsuitable.

The I.F. transformer must also be of the type tuned by an iron dust core at each end.

Dismantle the transformer, remove the parallel capacitors and mounting wires so that you are left with two coil bobbins and their former. Be very careful when un-soldering the wires from the supports because the coils are usually wound in litzendraht and must remain intact. Remove 25% (approximately) of the turns from one coil which we will call the secondary, and completely remove the iron dust slug from this coil, now cut the former in half, and mount as in Fig. 23a. Warm the coils with heat from your soldering iron or from a fire and gently slide the coils into position at the ends of the former as in Fig. 23a. The wax will set again holding the coils in their new position.

Note that the wires from the coils are connected to tags on the mountings. These are pieces of bakelite, and usually the original end pieces used on the I.F. transformer can be used.

A word about soldering litzendraht. Do not attempt to use emery paper. Prepare the end of wire by removing the outer silk covering and dip the end in methylated spirits. Ignite with a match and after about five seconds wipe with a cloth held between finger and thumb. This will quench the flame and will remove all traces of enamel from the wire. It is a good plan to experiment with the odd length of wire removed from the secondary coil. If the coils prove to be constructed of ordinary copper wire, it can of course be prepared with fine emery cloth in the normal manner.

Components List, Fig. 23a

C1 500pF variable capacitor.
X1 Germanium Crystal.
L1 See Text.
4 Plugs and sockets or terminals.

Wiring is straight forward as can be seen from Fig. 23b and requires no explanation.

Operation is as follows, selectivity is controlled by varying the distance between coils with the wing nut, and to an extent by the position of the primary iron dust core. Remember that the aerial and earth installations are as important with this set as with any other crystal receiver. No parallel capacitance appears necessary across the 'phones, the set works quite happily relying on the self-capacitance of the head-phones.

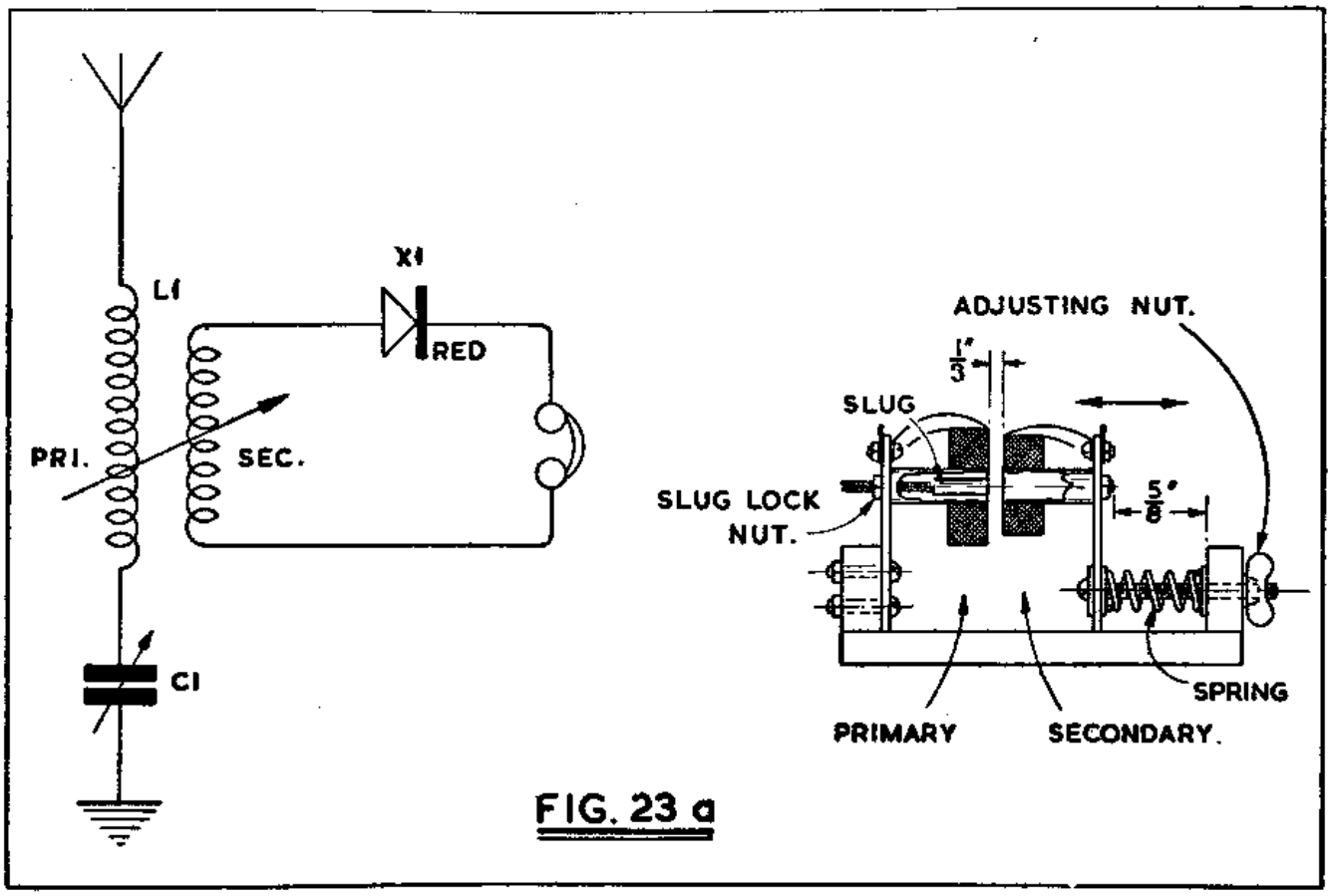


FIG. 23 a

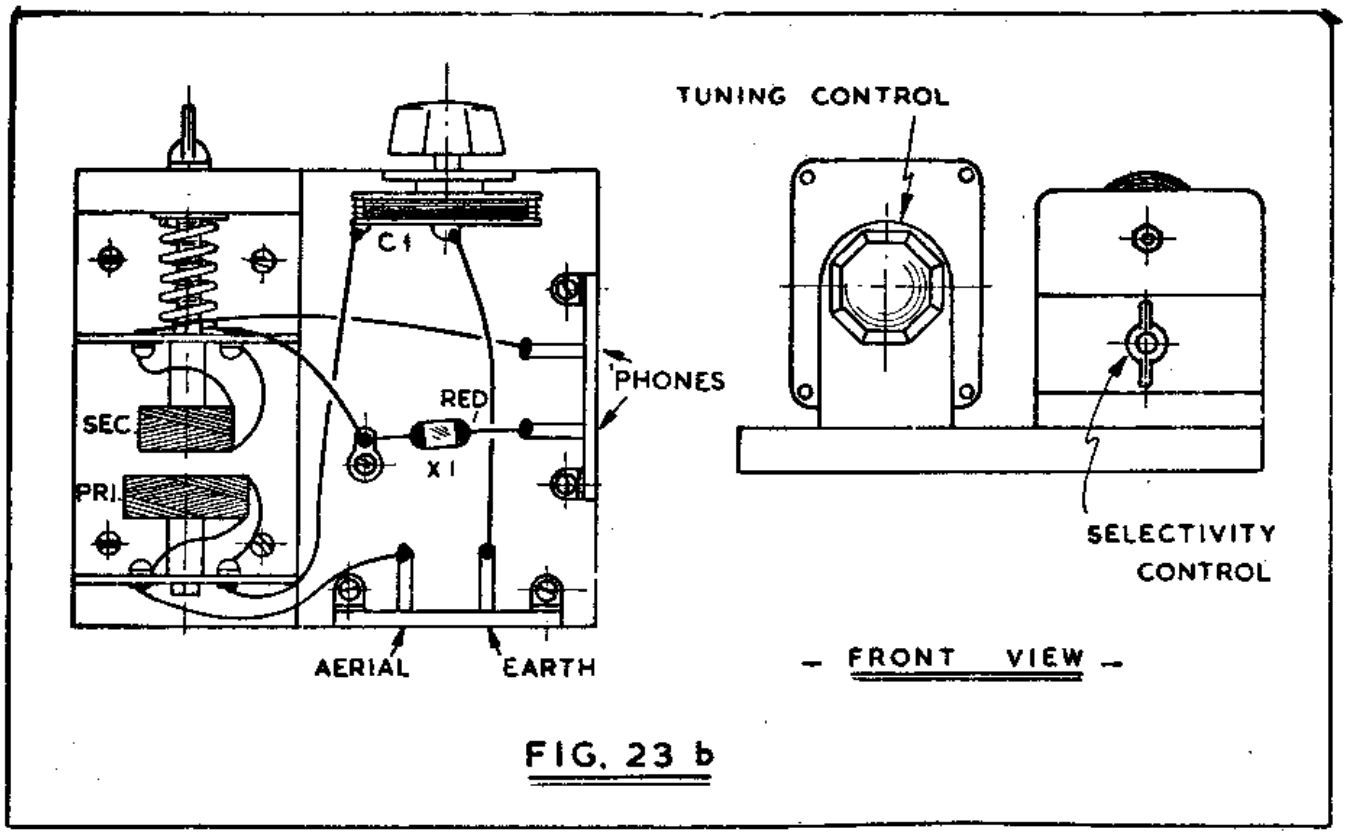


FIG. 23 b

9

So as to give a truly varied selection of circuits, some of the sets have been designed around commercial coils whilst others use coils which must be home-constructed.

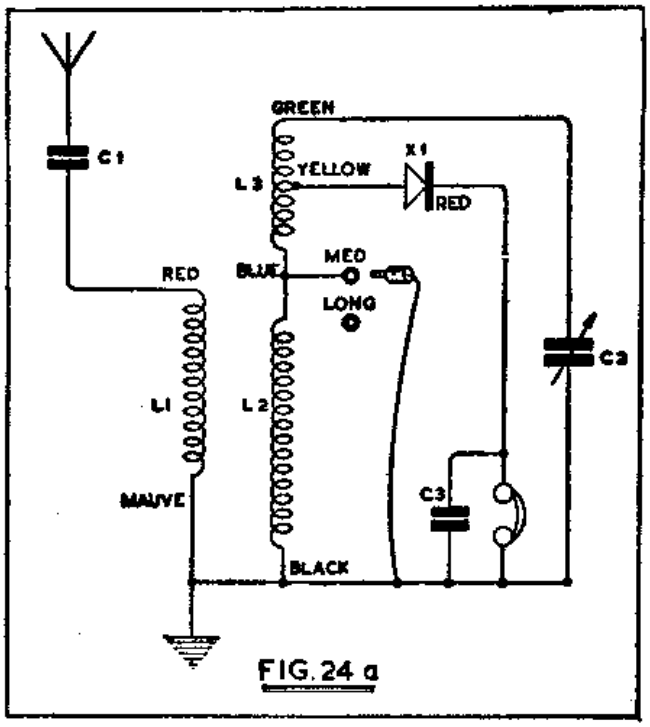


FIG. 24 a

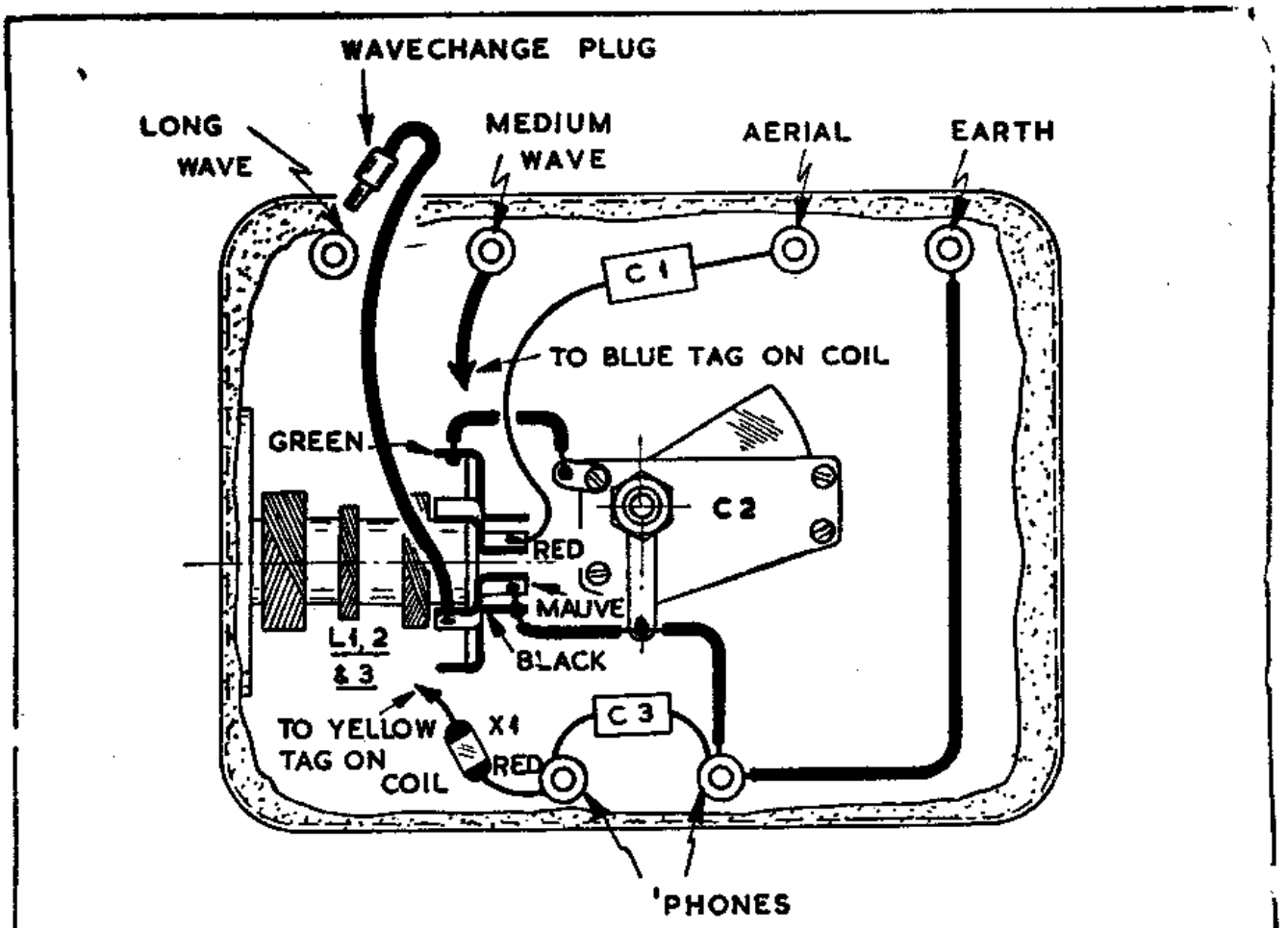
Most enthusiasts like to build their own coils but when it is a question of something really small combined with efficiency a commercial coil is the obvious choice.

Fig. 24a-b uses a tobacco tin or a similar container just as the design shown in Fig. 16a.

This receiver however is a little more selective and is more suitable than the earlier design, if you are near a transmitter or have a very long aerial. Take great care to connect the coil according to the colours shown on the diagrams. Sockets are used for wave-changing, this cheapens the cost a little and they are just as efficient.

Components List, Fig. 24a

C1 100pF Mica capacitor.
C2 500pF variable capacitor solid dielectric.
C3 1000pF Mica capacitor.
X1 Germanium crystal.
L1-2 R.E.P. dual range coil. Blue box.
6 Insulated wander-plug sockets.
5 Wander Plugs.
2oz. Tobacco tin or similar container.



LID SHOWN BROKEN AWAY TO REVEAL POSITION OF COMPONENTS.

FIG. 24 b

10

In certain cases, signals from the local transmitter are too powerful to permit a crystal set with a normal coil to be of much use. I myself live in Hertfordshire a few miles from the London transmitters, and find a more elaborate circuit is necessary. Now in radio circles it is a well-known fact that the greater the number of tuned circuits, the greater the overall selectivity. Look at Fig. 25a, you will see that the usual L and C tuning arrangements have been duplicated. This type of circuit is known as a Band-Pass Filter. There are many kinds of such filters, this particular type is an inductively-coupled filter, note that energy from the first half (L2) is conveyed to L3 by way of two small inductors L1 and L4.

Tuning is quite sharp and the "feel" of the set is quite different from that of single coil designs. A small chassis is ideal for construction.

6"x4"x2½" was used on the original, though it could have been smaller.

Fig. 25b gives the layout and wiring.

Take particular notice that L1-2-3 is mounted horizontally and at right angles to L4-5-6, this is to prevent the coupling from being too "tight" which would ruin the idea of the filter. A good aerial and earth is of course necessary. It may be found on test that the selectivity is too great

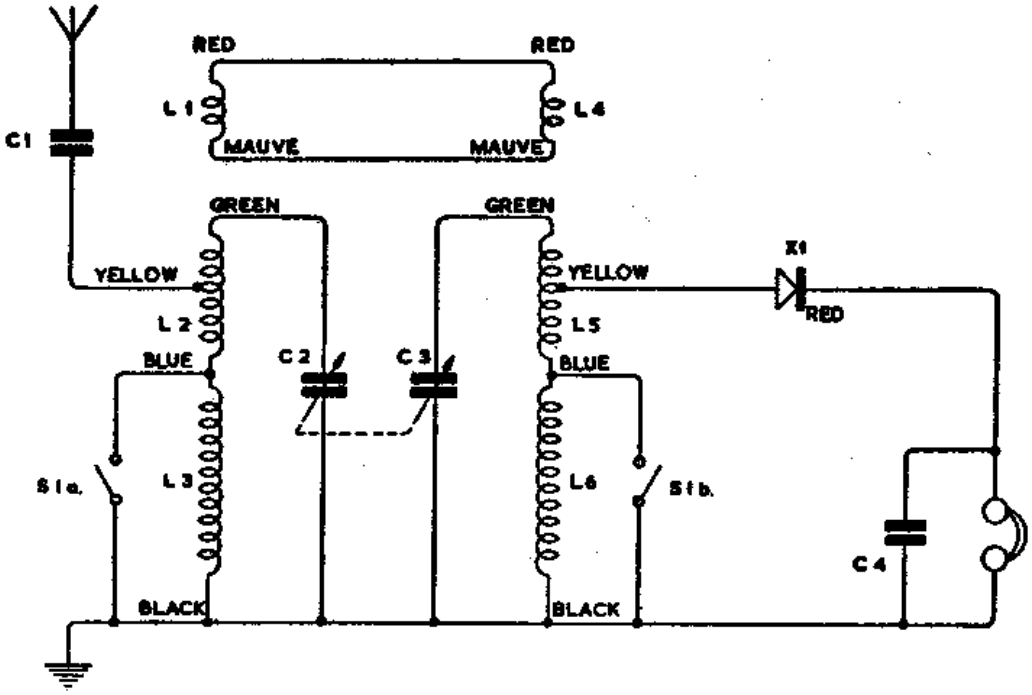


FIG. 25 a

and that volume has suffered. This can be ingeniously overcome by connecting a very small capacitance between the two Green tags on the coils. The simplest way is to connect it across the fixed vane connections on C2-C3. Values for this extra capacitance are a matter for experiment usually a value between 10pF and 47pF is sufficient. A small 50pF trimmer could be used and adjusted to suit. The ideal setting is to enable powerful adjacent stations to be received just short of overlap so that as much volume as possible is obtained.

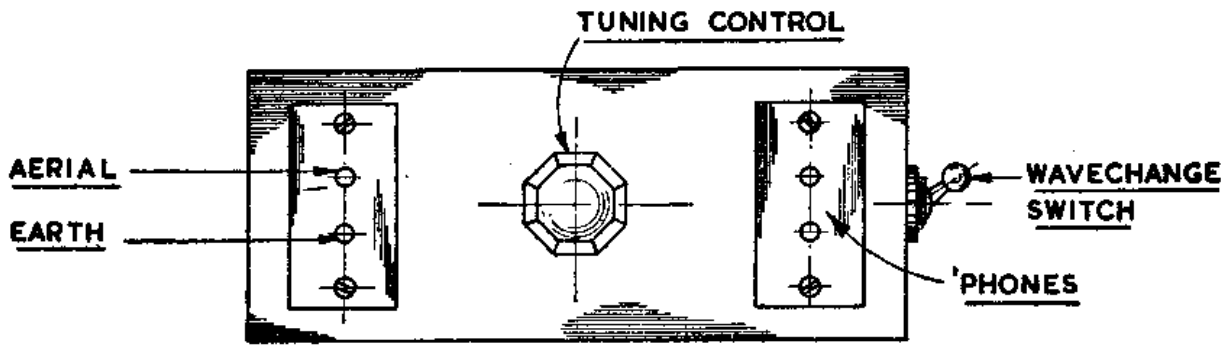
There is one point to be observed with the ganged capacitor. It should be of the type fitted with trimmers though of course, there is no reason why you should not fit them yourself. These are to allow for differences between coils to be balanced out, and consist of a small variable capacitance across each main section.

When setting up the receiver, first set each trimmer at half way and tune in a station near the high-frequency end of the band (vanes nearly out) adjust the trimmers for maximum volume. If extra coupling capacity is added as described, try adjusting them further.

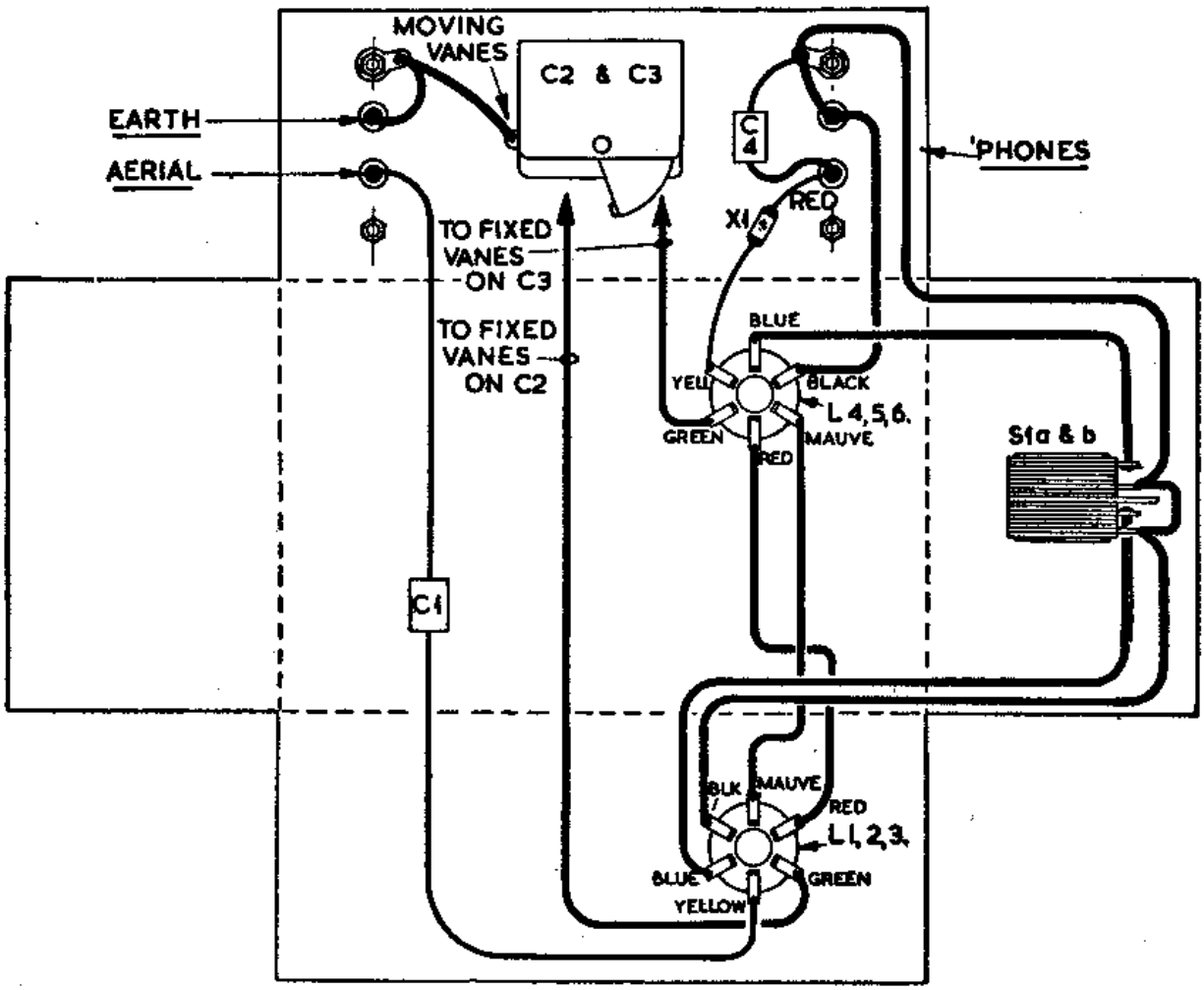
Once properly set up, no further adjustment is necessary.

Components List, Fig. 25a

C1 100pF Mica capacitor.
C2-3 2 x 500pF variable capacitor (see text).
C4 1000pF Mica capacitor.
L1-2-3 Dual Range Coil R.E.P. (Blue Box).
L4-5-6 Dual Range Coil R.E.P. (Blue Box).
X1 Germanium Crystal.
S1a-b 2 pole single throw toggle switch.
4 Terminals or plugs and sockets.



FRONT OF COMPLETED RECEIVER.



EXPLODED UNDER-CHASSIS VIEW

FIG. 25 b

11

Another band-pass circuit is shown in Fig. 26a. This time home-made coils are used. This filter is not inductively coupled but entirely capacitive, relying on C3. As in the previous circuit it must be adjusted to give just sufficient " broadness " to the tuning to provide adequate volume without station overlap. To avoid unwanted coupling a screen is mounted between the coils, this is necessary because due to the size of the coils they would be bound to couple quite tightly without it. Such precautions were not necessary with the circuit of Fig. 25a, as the commercial coils are much smaller physically and the risk of unwanted coupling reduced.

The coils for this receiver are for medium wave reception only, this keeps the size down. Coil winding procedure has already been fully explained, and covers all the requirements of these coils. Base-board and panel construction is used, and for more accurate tuning since the coils are home-made, separate tuning capacitors are recommended. Fig. 26b gives the layout and practical wiring.

Components List, Fig. 26a

C1 100pF Mica capacitor.
C2 500pF variable capacitor.
C3 50pF Mica trimmer.
C4 500pF variable capacitor.
C5 1000pF mica capacitor.
L1-2 See text.
X1 Germanium Crystal.
4 Terminals or plugs and sockets.

For those who do not wish to bother with these a ganged capacitor can be used, but tuning may not be as good as with separate ones. The screen mounted between the coils is of aluminium or copper and should be about 1" higher than the coils and the same length as the baseboard depth.

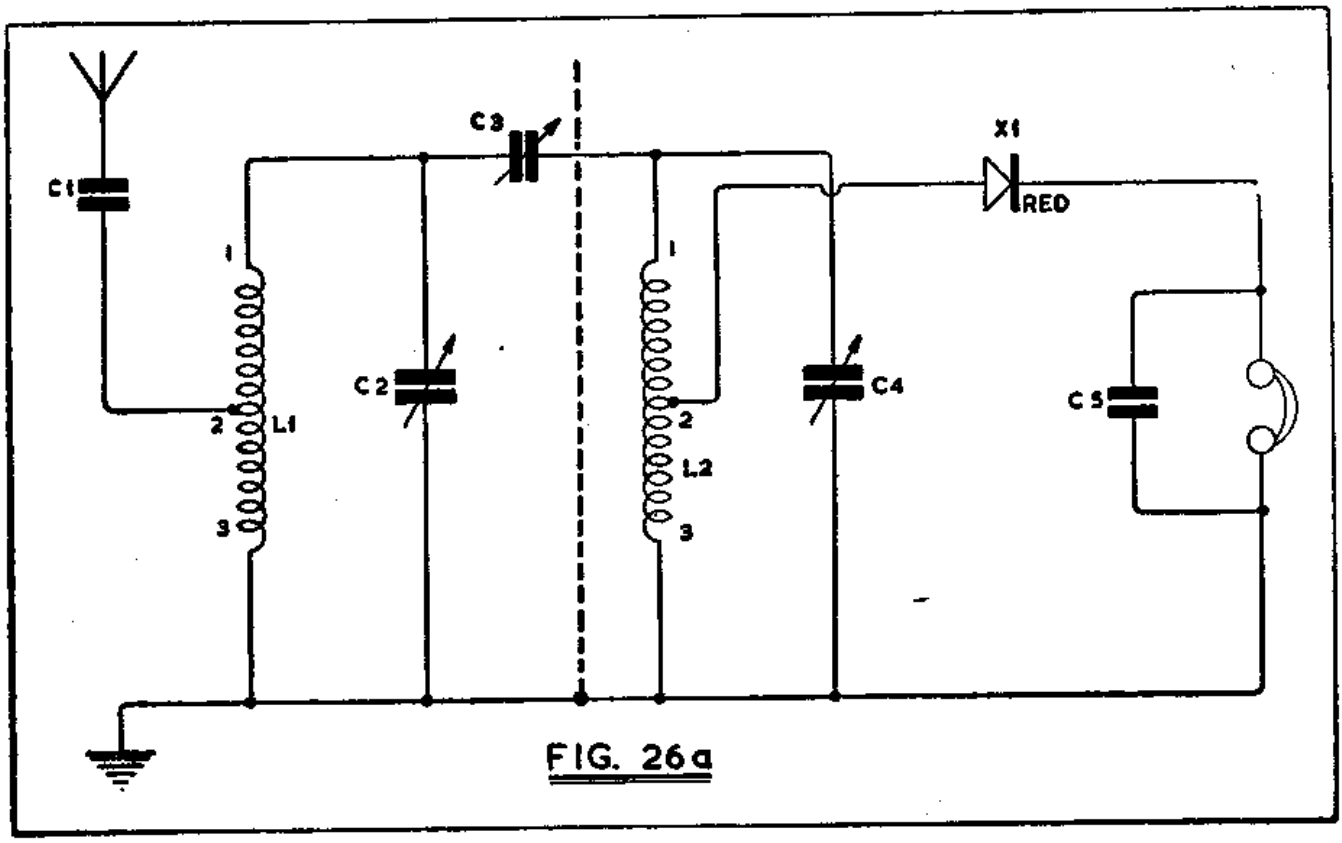
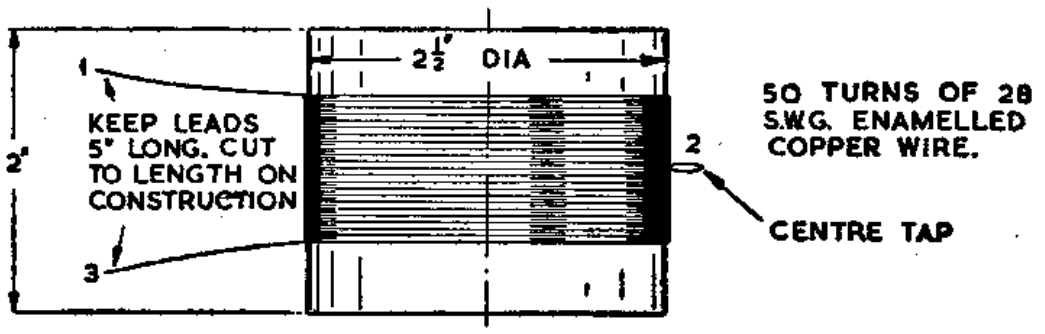


FIG. 26a



COIL DETAILS.

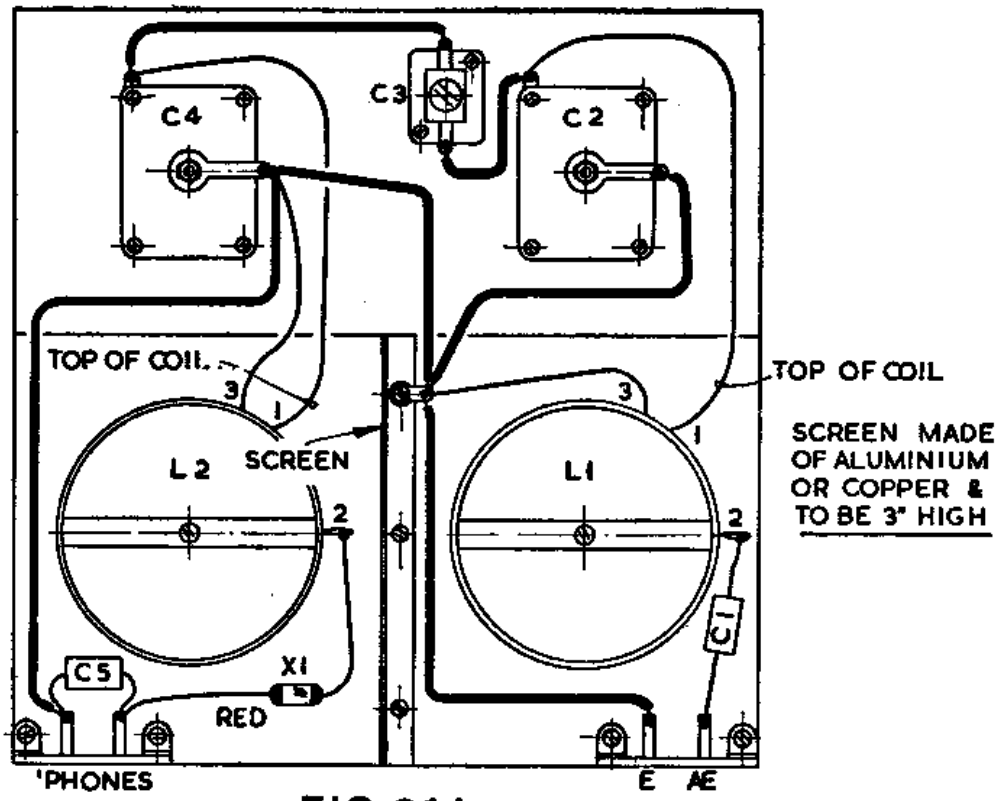


FIG. 26 b

12

Fig. 27a-b uses a band-pass filter for tuning, and the coils are home-constructed.

With this design a ganged capacitor is quite suitable as the coils tune quite accurately. It is essential that the coupling condenser C3 is **non-inductive,** there will be no difficulty in obtaining a new component of this pattern but it is as well to avoid old components which may not employ this form of construction.

Receivers using this type of filter are used to the best advantage when situated a short distance from a powerful transmitter and if selectivity is not important this particular design is not the most suitable.

Both medium-waves and long-waves are catered for: consider L1-2-3, this is the first half of the filter.

L1 is the aerial coupling coil, with S1 closed,

the medium-waves are received. This coil is inductively coupled to L2 and L3. Again with S2 open L2 and L3 combined cover the long-waves and when S2 is closed the medium waves are covered.

Tuning for this half of the filter is by means of the capacitor C1, which is one half of the ganged capacitor. A screen is erected between the two sets of coils comprising the filter, and coupling is effected by C3, an earth return for the coils is provided by R1.

L4 and L5 operated in a similar manner to L2 and L3, tuning is by means of C2, the remaining section of the ganged capacitor. You will realise that the coils must be accurately wound otherwise tuning will not remain constant between the two sections over the band.

The ganged capacitor should be fitted with trimmer capacitors when purchased so that the two circuits can be balanced.

You will have noticed that this circuit uses three sets of switch contacts. A three-pole two-way wafer switch is the best choice otherwise either two or even three separate switches are necessary. All the necessary coil details are given in the diagrams, the tap on L4 is made in the same way as shown for other coils in the manual. One other component used only on this design, is the resistor R1, its size is not important because the current flowing is so small it can be ignored.

A ½ watt rating is adequate, its precise value is not critical so that a 20% tolerance component is quite suitable.

Resistors are coded by colours and the 1000Ω type you require will bear the following sequence. At one end will be painted three coloured rings.

Components List, Fig. 27a

C1-2 500pF 2 gang, variable capacitor.
C3 0.05mfd. non-inductive paper capacitor.
C4 1000pF mica capacitor.
R1 1000 ohm resistor.
X1 Germanium Crystal.
S1-2-3 3-pole 2-way wafer switch.
L1-2-3 & L4-5 See text.
4 Terminals (Aerial, Earth and 'Phones).
Sheet copper or aluminium for screen.

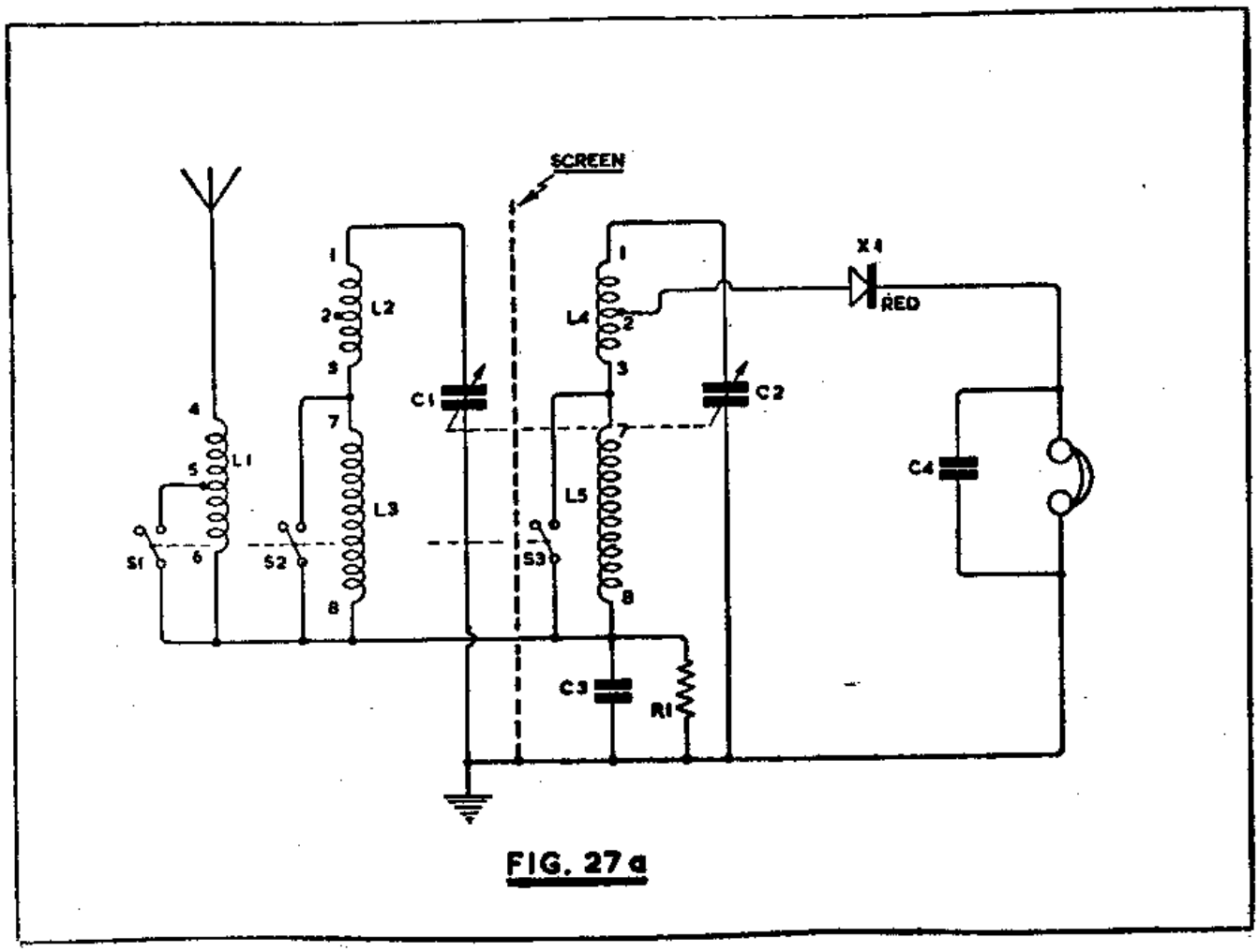
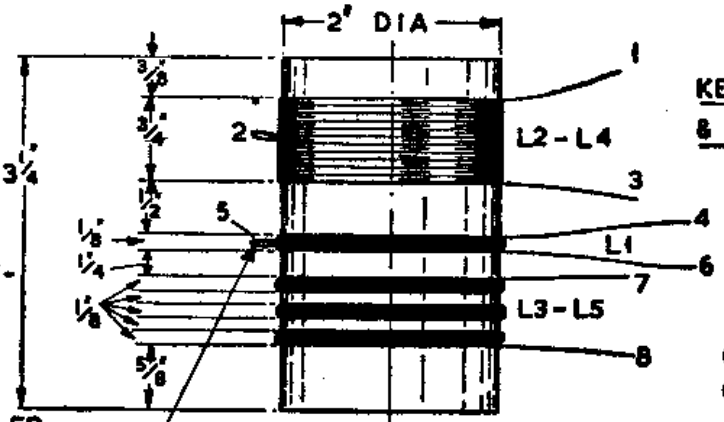


FIG. 27a

L2 - L4
58 TURNS No.30 S.W.G. ENAMELLED WIRE. CENTRE TAPPED.

L1. 75 TURNS No.36 S.W.G. ENAMELLED WIRE - PILE WOUND.

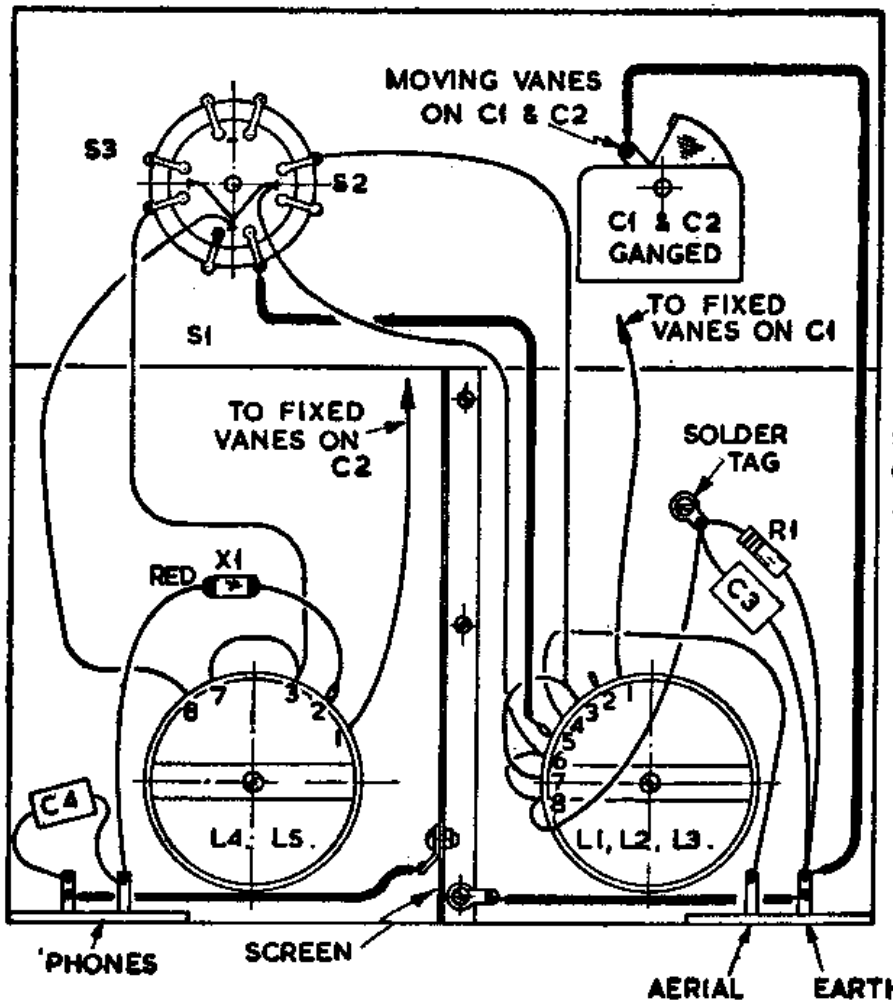
L3 - L5
174 TURNS No. 36 S.W.G. ENAMELLED WIRE - WOUND IN 3 PILES OF 58 TURNS



COIL DETAILS.

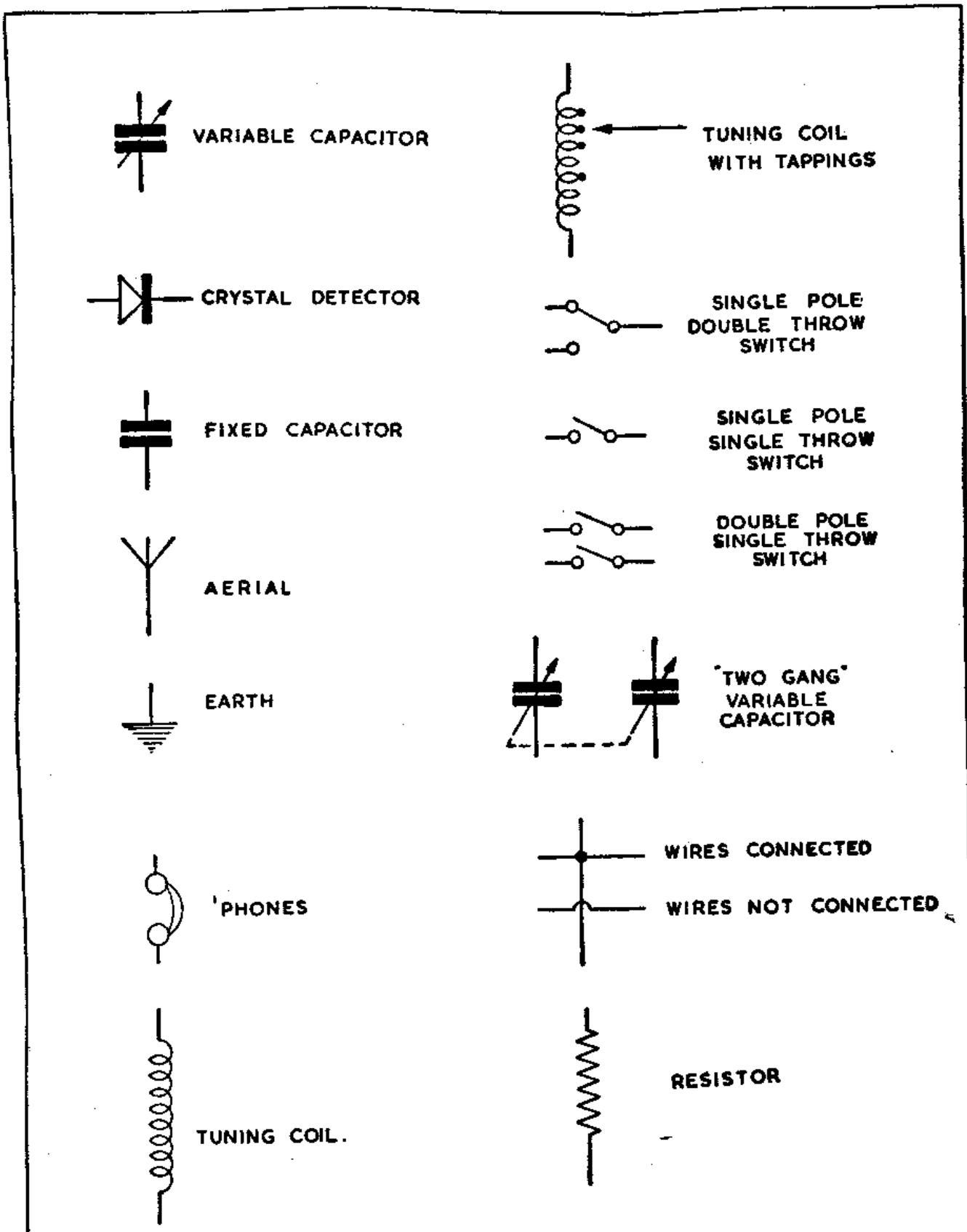
KEEP LEADS 6" LONG & CUT TO LENGTH ON CONSTRUCTION

ON COIL L4 & L5, OMIT WINDINGS 4 TO 6



SCREEN MADE OF COPPER OR ALUMINIUM & TO BE 4½" HIGH.

FIG. 27 b



SYMBOLS USED IN THE CIRCUITS DESCRIBED.

FIG. 28

The first one is BROWN which represents 1, the first figure of the value, the second BLACK, this indicates the second numeral is nought, and the third colour is RED which tells us that two further ciphers or noughts are added to give the total value. From this you can see, we have 1 plus a nought equalling 10, plus two further noughts which total 1000. Some earlier types of resistor were painted differently, the whole body colour represented the first figure, one tip was coloured to represent the second figure, and a painted dot in the centre gave the number of ciphers to be added.

There are no other special points regarding construction except perhaps that if a long aerial is used, a small 100pF capacitor might be tried in the aerial lead to prevent damping the first tuned circuit.

When setting up the receiver, tune to a station the high-frequency end of the M.W. band and the trimmers for loudest volume, no other adjustments are necessary.

CONCLUSION

The twelve receivers described are representative of the best crystal set designs available to-day. Their construction will provide many hours of useful enjoyment, and the results will give lasting pleasure.

A final word about components, if no actual maker is specified, any good class component can be used. Switches, fixed and variable capacitors and crystals, are available from numerous manufacturers all of which are invariably of excellent quality. Where a particular manufacturer's product is called for, the specification should be adhered to. The commercial coils specified are freely available from most supply houses specialising in components for constructors, but in case of difficulty write to Bernards (Publishers) Ltd., The Grampians, Western Gate, London, W.6, who will, on receipt of a S.A.E. be pleased to supply the address of your nearest stockist. Coil formers are not always easy to obtain and in this case advice should be sought from Post Radio Supplies, 395, Queensbridge Road, London, E.8, who will be pleased to help in any way possible.

BERNARDS RADIO BOOKS

No.		Price
56.	Radio Aerial Handbook	2/6
57.	Ultra-Shortwave Handbook	2/6
58.	Radio Hints Manual	2/6
64.	Sound Equipment Manual	2/6
68.	Frequency Modulation Receivers' Manual	2/6
83.	Radio Instruments and their Construction	2/6
96.	Crystal Set Construction	1/-
99.	One Valve Receivers	1/6
100.	A Comprehensive Radio Valve Guide, Book 1	5/-
103.	"Radiofolder" A. The Master Colour Code Index for Radio and T.V.	1/6
104.	Three Valve Receivers	1/6
121.	A Comprehensive Radio Valve Guide, Book 2	5/-
123.	"Radiofolder" F. The Beginners' Push-Pull Amplifier	1/6
126.	Boys' Book of Crystal Sets and Simple Circuits	2/6
129.	Universal Gram Motor Speed Indicator	1/-
135.	All Dry Battery Portable Construction	2/6
138.	How to Make F.M. and T.V. Aerials, Bands 1, 2 and 3	2/6
141.	Radio Servicing for Amateurs	3/6
143.	A Comprehensive Radio Valve Guide, Book 3	5/-
145.	Handbook of AM/FM Circuits and Components	2/-
146.	High Fidelity Loudspeaker Enclosures	5/-
147.	Practical Tape Recording Handbook	5/-
148.	Practical Transistor Receivers, Book 1	5/-
149.	Practical Stereo Handbook	3/6
150.	Practical Radio Inside Out	3/6
151.	Transistor Superhet Receivers	7/6
156.	Transistor Circuits Manual, No. 1	2/6
157.	A Comprehensive Radio Valve Guide, Book 4	5/-
158.	Radio, Television, Industrial Tubes, Semiconductor and Diodes Equivalents Handbook (208 pages)	9/6
159.	Realistic High Fidelity	5/-
160.	Coil Design and Construction Manual	5/-
161.	Radio, Television and Electronics Data Manual	3/6
162.	High Fidelity Stereo Gramophone	5/-
163.	Transistor Circuits Manual, No. 2	2/6
165.	Hartley on Hi-Fi, Book 1. Radio Tuners	5/-
166.	Public Address Systems	2/6
168.	Transistor Circuits Manual, No. 4	2/6
170.	Transistor Circuits for Radio Controlled Models	7/6
171.	Super Sensitive Transistorised Pocket Set Construction	3/6
172.	International Radio Stations List	2/6
173.	Practical Transistor Audio Amplifiers	3/6
174.	Transistor Subminiature Receivers Handbook	5/-
175.	Transistorised Test Equipment and Servicing Manual	3/6
176.	Transistor Audio Amplifiers Manual	6/-
177.	Modern Transistor Circuits for Beginners	7/6
178.	Comprehensive Radio Valve Guide Book 5	6/-
179.	Transistor Circuits Manual No. 5	5/-
180.	British Semiconductor Survey	15/-
181.	22 Tested Circuits Using Micro Alloy Transistors	5/-
182.	"At A Glance" Radio Valve & T.V. Tube Equivalents	3/6
183.	How to Receive Foreign T.V. Programmes on your Television Set by simple Modifications	5/-
184.	Tested Transistor Circuits Handbook	2/6
185.	Tested Short-Wave Receiver Circuits using Micro Alloy Transistors (MAT's)	5/-
186.	Tested Superhet Circuits for Short-Wave and Communication Receivers using Micro Alloy Transistors (MAT's)	6/-
187.	The TSL Mark "4" Valved F.M. Tuner and its Construction	3/6
188.	Construction of the B.H. High Fidelity "14" watt Audio Amplifier	1/6
189.	How to build the world's smallest transistor radio. The Sinclair Micro 6	1/-
190.	How to build the world's smallest High Fidelity Amplifier	1/6
191.	Practical Car Radio Handbook	6/-
501.	ABC's of Magnetism	5/-
502.	ABC's of Missile Guidance	5/-
503.	ABC's of Ultrasonics	5/-
504.	ABC's of Transistors	5/-
	Resistor Colour Code Disc Calculator	1/6
	Engineers' Reference Tables	1/6
	International Radio Tube Encyclopaedia—3rd Edition	63/-

Build a 3-Band Shortwave Converter for Your Car

Wouldn't it be nice if you could take your shortwave radio with you in your car? Imagine tooling down the highway on your way to work, listening to the latest world news on the BBC. Or maybe tuning in a spot of upbeat Latin music to get your day started on the right foot.

You could put an inexpensive portable on the dashboard and watch it slide from one side to the other every time you make a turn. The heat of the sun, magnified through the windshield, even in winter, might ruin that radio and chances are that if it survives that, someone might steal it. Too much of a headache? There is an easier way.

We're going to show you how to build a converter for your car radio that will allow you to tune in the 49 (5950 to 6200 kHz), 31 (9500 to 10000 kHz) and 19 (15100 to 15490 kHz) meter bands without all of those hassles. We chose these bands because the probability of finding one or more of them open during any time of the day or night is good. Also, the time signals from WWV on 10000 kHz are tuneable on the 31 meter band. So we have a talking clock in our car to tell us how late we are!

How it Works

There are two controls on the converter; the bandswitch S1, and the on/off switch S2. T1 matches the car antenna's impedance to the input impedance of Q1, a dual gate mosfet. Capacitors are placed in parallel with T1's secondary by S1A. These capacitors are chosen to resonate T1 to the band we wish to tune on the car radio.

Q1 is the mixer. Shortwave band frequencies are present on gate #1, while a fixed H.F. frequency from oscillator Q2 is present on gate #2. At the drain of the mixer are 4 signals: the 2 original ones on gates 1 and 2, their sum, and their difference. The difference frequency will fall in the tuning range of the AM car radio.

Q2 is a Pierce crystal oscillator. Two capacitors (designated Cfb) provide feedback to sustain oscillation. S1B selects a crystal to correspond to the band selected by S1A. The oscillator is a fundamental type, meaning it's output is the same frequency as the crystal.

S2 is a DPDT slide switch. One pole switches 12v DC to the converter during use, the other pole switches the converter out of the antenna line when not in use.

Choosing Crystals

The builder is given wide latitude in choosing crystals for this converter, with the hopeful result that he will be able to scrounge them up for free in the junkbox or for cheap at a swapmeet or hamfest. It makes no difference if the crystal frequency is above or below the shortwave frequency we want to receive, so long as the difference falls in the tuning range of the AM car radio.

If the crystal frequency is below the shortwave frequency, then the car radio will tune the shortwave band in the conventional manner (lowest frequency on the left, highest on the right of the slide rule scale). If the crystal frequency is above the shortwave frequency, then the car radio will tune "backwards" (highest frequency left, lowest on the right).

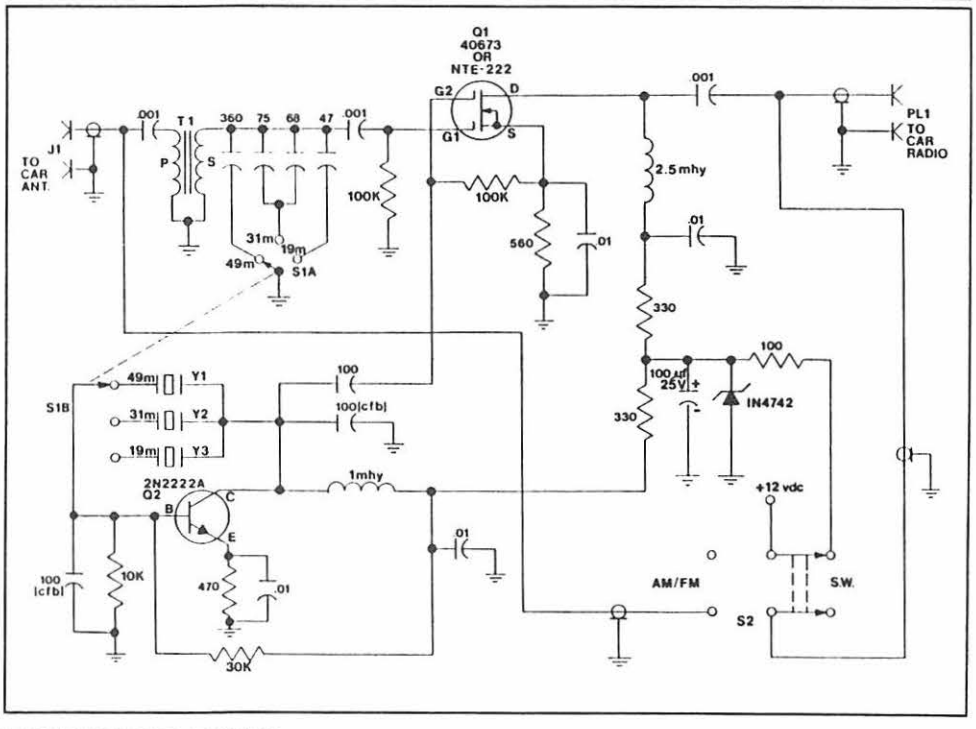
Listening enjoyment will not be affected if you choose a crystal which causes "backward" tuning. So start scrounging for crystals that will fall in the frequency ranges given in the chart below, and see how cheaply you can get by!

Construction Tips

The manner of construction is not too critical. Perfboard is recommended for the first time builder. An ambitious builder could make up a single or double sided PCB. T1 must be wound by hand; the secondary windings must use the gauge of wire specified and must be "closewound" in order for the transformer to resonate properly with the values of fixed capacitors specified in the schematic. A sketch of how the transformer should look is included with the semiconductor base diagrams.

Crystal Frequency Ranges

Xtal	Band	Band Frequency	Xtal Frequency Range
Y1	49M	5.95-6.2 MHz	4.6-5.41 MHz or 6.74-7.55 MHz
Y2	31M	9.5-10.0 MHz	8.4-8.96 MHz or 10.54-11.1 MHz
Y3	19M	15.1-15.49 MHz	13.895-14.56 MHz or 16.035-16.7



Keep all leads to S1 as short as possible. One way to do this is to mount S1 on the front panel of the box you are going to use, then mount your board directly behind S1 using screws, spacers, and nuts. (see diagram). J1 and PL1 are simply a 2 foot long car antenna coax extension cable available from Radio Shack, which is cut in half. Drill two holes in the back of the box and fit the holes with grommets, then put the two pieces of coax through the holes. The +12v power lead can also pass through one of the holes with the coax.

Be sure and use a metal box for the enclosure, as a plastic one will allow the strong local mediumwave stations to "bleed" past the mixer and you'll never hear any shortwave stations. The fixed capacitors on S1A could be replaced by small trimmer capacitors of a slightly higher value if desired. This will increase the costs but will allow you to more exactly resonate T1 to the center of each shortwave band.

Using the Converter

First of all be advised that this project will work best with a conventional slide-rule tuned (analog) car radio. Most digital radios are pre-programmed to tune in 10 khz increments, where as shortwave stations are in 5 khz increments or on "weird" frequencies like 99770 kHz. So a digital car radio which is programmed in 10 khz increments will only allow you to hear half of what is available.

Next you need to know where on your dial you can expect to find the shortwave. To do this, subtract the frequency of each crystal you've found from the band limits of it's corresponding band. For example, say you found a crystal marked 8.65 MHz for the 31 meter band. 9.5 - 8.65 = .85, and 10.0 - 8.65 = 1.35. So you will tune the 31 meter band between 850 and 1350 on your car radio, with WWV falling on 1350.

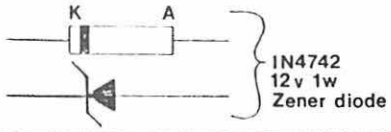
This circuit is sensitive enough to get all the major broadcasters loud and clear with your engine on. And if you are out camping in a quiet location with your engine off, you can actually do some real DXing on your car radio using only the car's antenna!

Good luck! Any questions from constructors regarding this project will be answered if a Self-addressed, stamped envelope is enclosed.

Eric Johnson KB6EPO
799 Ada Street
Chula Vista, CA 92011-2603

SEMICONDUCTORS

DEVICE	BOTTOM VIEW	TOP VIEW
40673 OR NTE-222 N-CHANNEL dual gate MOSFET	G2, G1, D, S	G2, G2, S, D
2N2222 NPN Bipolar Transistor	C, B, E	C, B, E

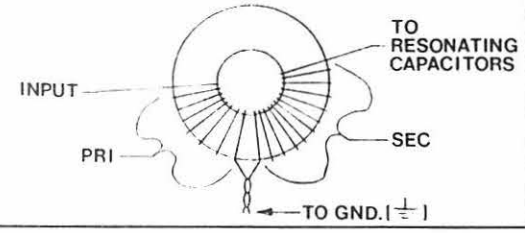


TOROIDAL TRANSFORMER T1

AMIDON T-68-2 core (red color, one size.)

PRI: 8 Turns, #21 enamel coated wire.

SEC: 12 Turns, #21 enamel coated wire.

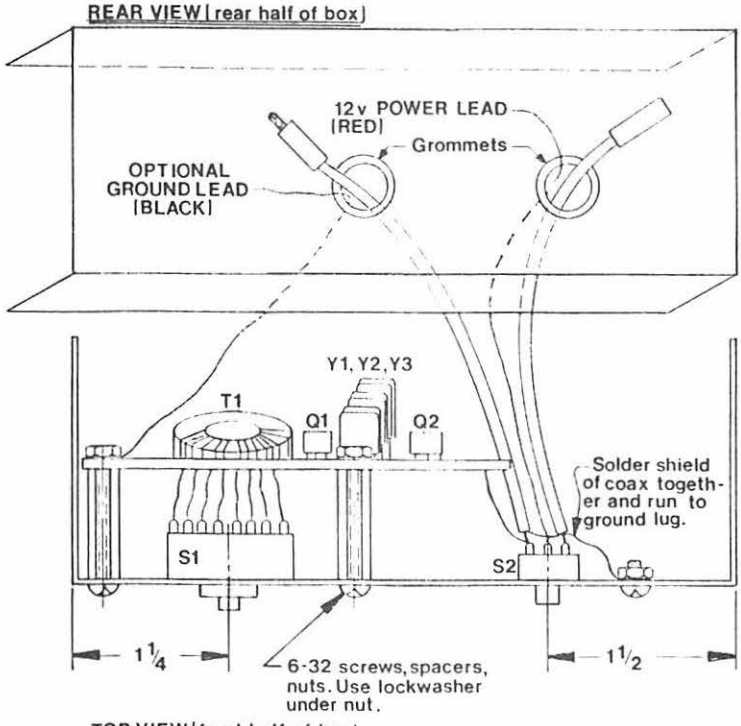


GENERALIZED PHYSICAL LAYOUT

1.) Use metal box 5 1/4" W x 2 1/4" H x 2" D; larger if you feel confined working in so small a box. Box consists of two U shaped pieces held together by screws.

2.) Mount back half of box to frame of car with screws. If mounted to plastic, provide ground wire as shown, connect to car chassis.

REAR VIEW (rear half of box)



TOP VIEW (front half of box)

* Hook power lead to "radio" side of fuse, or use an in-line fuse holder with a 1/4 A fuse inside.

Cigar Box Radio

BACK in the early days of radio, it was a common sight to see the entire family crowding around and taking their turn at the earphones of the latest marvel of the age—the crystal set radio. It seemed so uncanny how it was possible to get music and voices in the room from afar without the use of wires. With the advent of newer radios the crystal set slowly lost its popularity. However, it still is lots of fun to turn the pages of time back and build a simple crystal set. Anyone with but average workshop skill can easily construct such a set in an evening or two, and, if located about ten miles from a broadcasting station reception should be good.

Winding the coil. Coil tube, with its wooden center core, revolves when the handle is turned.

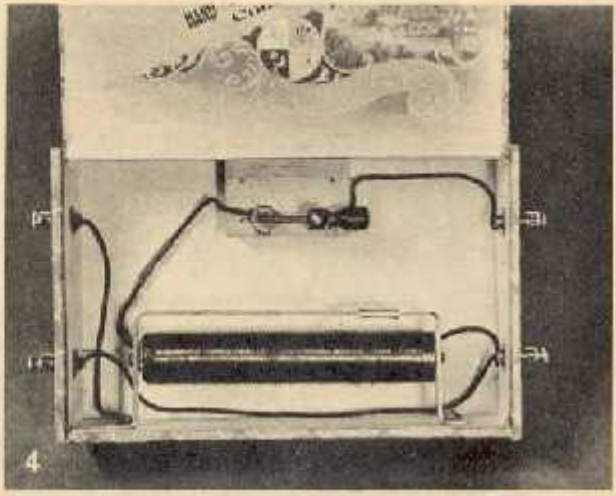
After the ends of the coil are made secure, apply a thin coat of shellac to the entire surface.

Carefully solder a wire lead to each end of the coil. Use No. 18 or 20 insulated flexible wire.

The first step is to make the tuning coil inductance that is used in the antenna circuit. Take a piece of ¾ to 1 in. diameter mailing tube or fibre tubing about 6 inches long and closely wind on a single layer of No. 26 to 28 gauge enameled wire. For winding the coil make the simple jig shown in Photo 1. The jig consists of a base board with two L-shaped metal brackets. The coil tube is set between the brackets with a screw holding one end and a piece of ⅛ in. steel rod which acts as a handle, imbedded into the other end. Both screw and handle rod are driven into a round piece of wood pressed into the tube for a center core. The wooden center is left in place and later permits mounting the finished coil permanently in the cigar box. The end of the steel rod, which forms the turning handle, is flattened where it is driven into the wood.

The start and finish of the coil are secured from unwinding by drilling a small hole at each end of the tube and passing a loop of the wire through each hole. The coil is then given a coat of thin shellac, as illustrated in Photo 2.

The next job, shown in Photo 3, is to



Arrangement of parts within the cigar box. Note the coil mount, slider, and crystal detector.

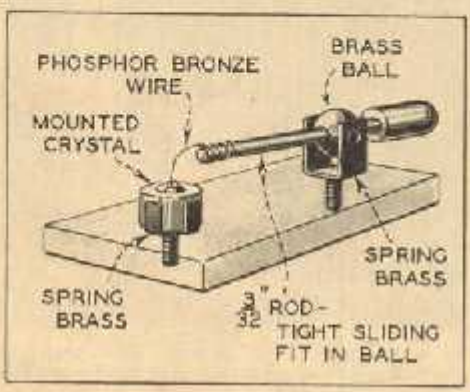
BY HAROLD P. STRAND

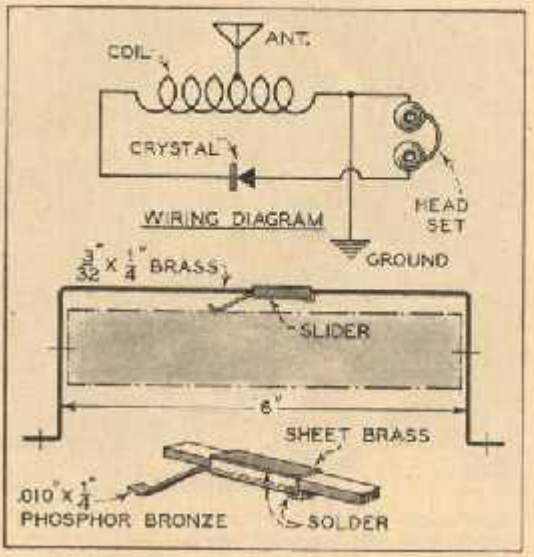
solder a piece of No. 18 or 20 insulated flexible wire to the start and finish of the coil to act as leads. Each wire lead is made secure by inserting the end through the hole drilled for the winding ends.

Next mount the coil to the front side of the box with short bolts inserted through a combination support and slider arrangement, as pictured in Photo 4. The support is merely a strip of $\frac{1}{4} \times \frac{3}{16}$ in. brass bent to fit the coil. The slider, built up from sheet brass, slides freely on the strip. Clean the coil over which the slider rides with fine sandpaper, down to the copper, so a good contact will be made with every turn. Wood screws are used at each end of the coil to hold it in place.

The detector is a common mounted galena crystal and an adjustable arm with a "cat whisker," obtainable from any radio or electrical store. At each end of the box binding posts are fitted. The upper one at the left side is for the antenna connection and the lower one for the ground. The right hand terminals are for the phones. The antenna comes in and connects to the brass slider at its bottom foot. The ground connects to one phone terminal and also to one end of the coil. The other end of the coil goes to the crystal and the detector arm to the remaining phone terminal.

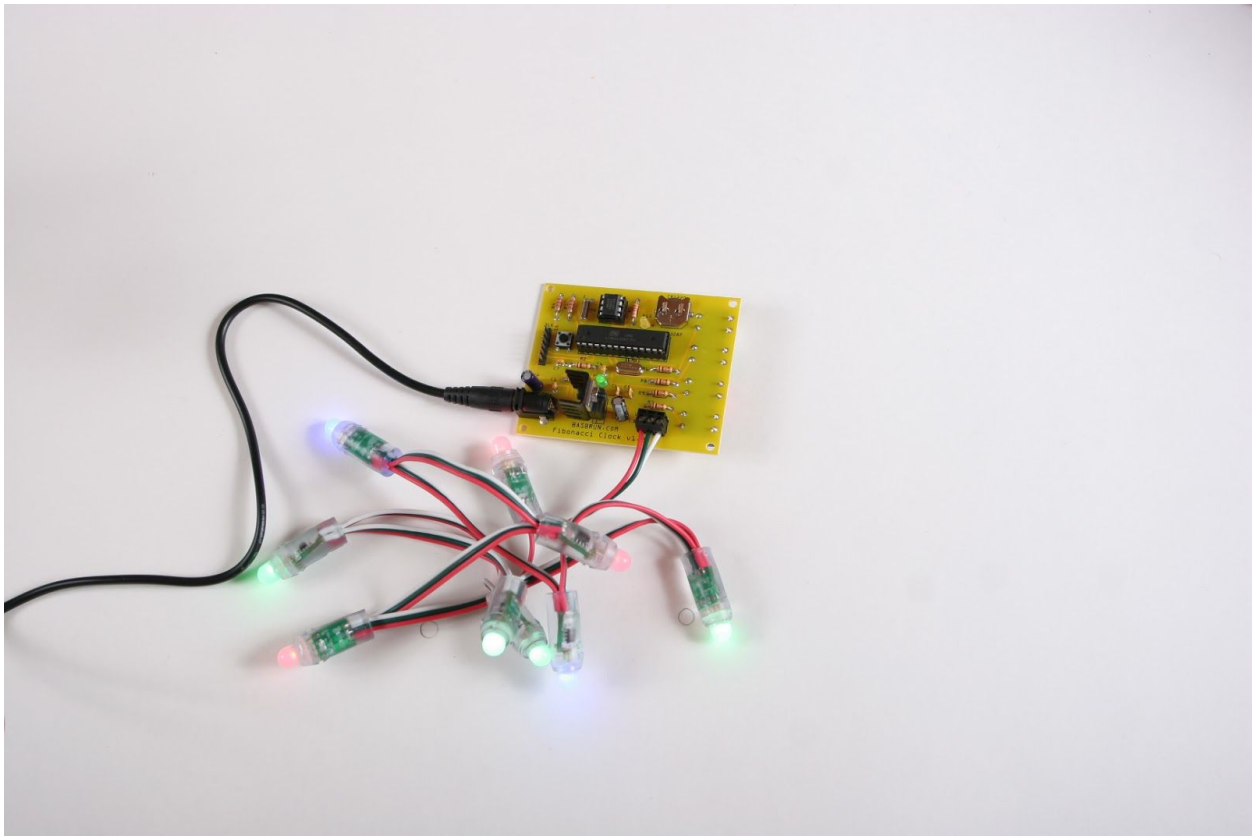
For best results a long outside antenna of about 75-100 ft. long is recommended. The earphones should be sensitive and wound to a fairly high resistance, such as 2000 ohms. Adjustment is provided by the sliding contact which varies the tuning inductance. It should be possible to get several local stations on different spots on the coil, but selectivity will be only fair so in some cases interference may be experienced. The detector must be very carefully adjusted to find a sensitive spot by moving the cat whisker around to touch different places on the crystal.





Circuit Board Assembly Instructions

This document walk you through the assembly of the Fibonacci Clock v1.2 circuit board.



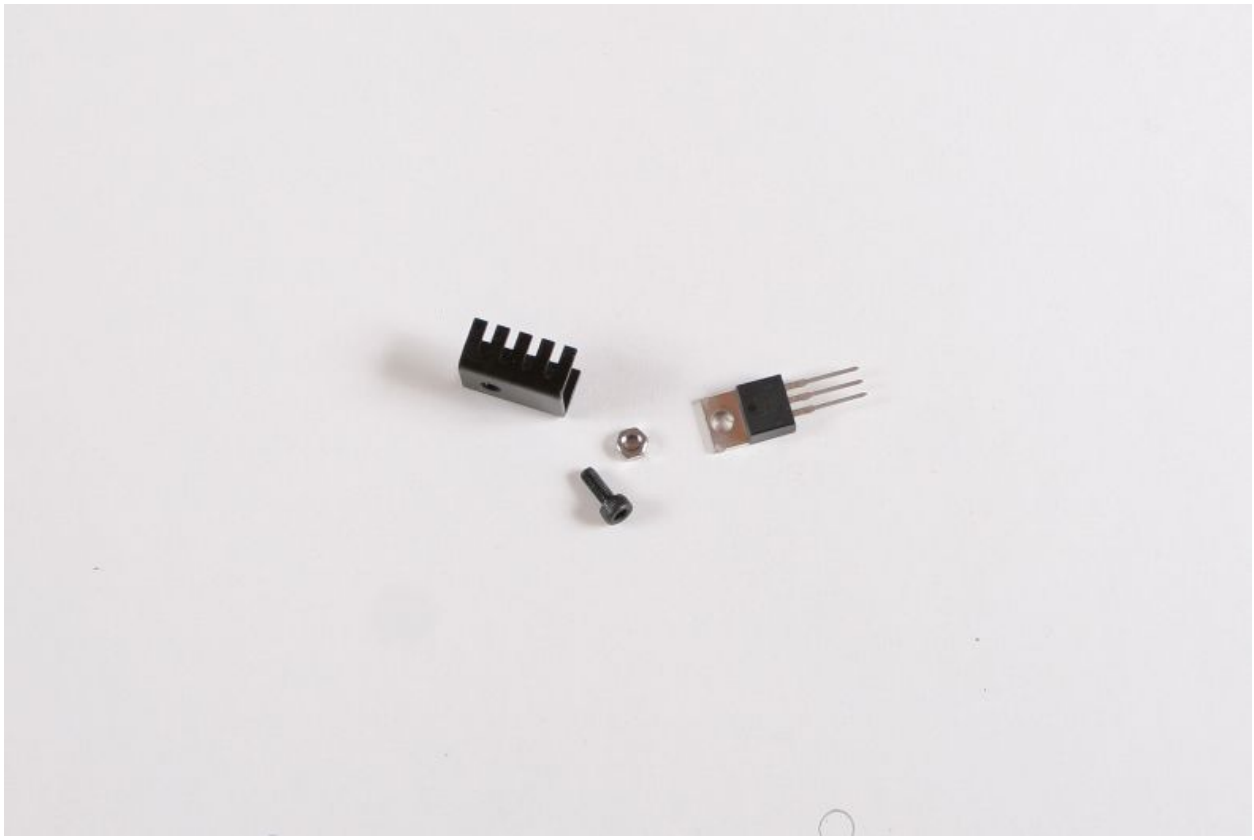
Important note for kit buyers **The color and appearance of the components may vary from one supplier to another and not match exactly the components illustrated in this document.**

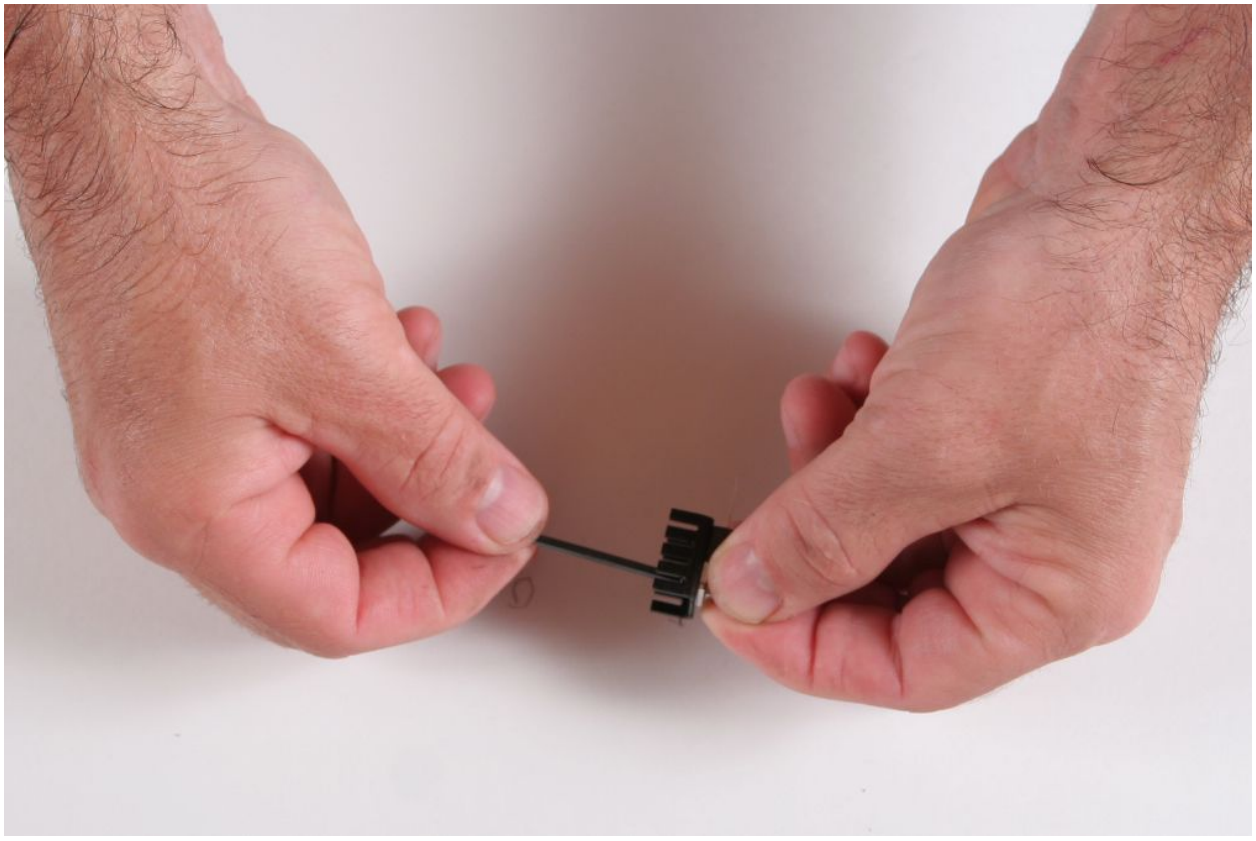
The kit provided for the Kickstarter backers matches the images here with the exception of the ceramic capacitors and one electrolytic capacitor. Some notes have been added to the “Parts” section when the appearance differ.

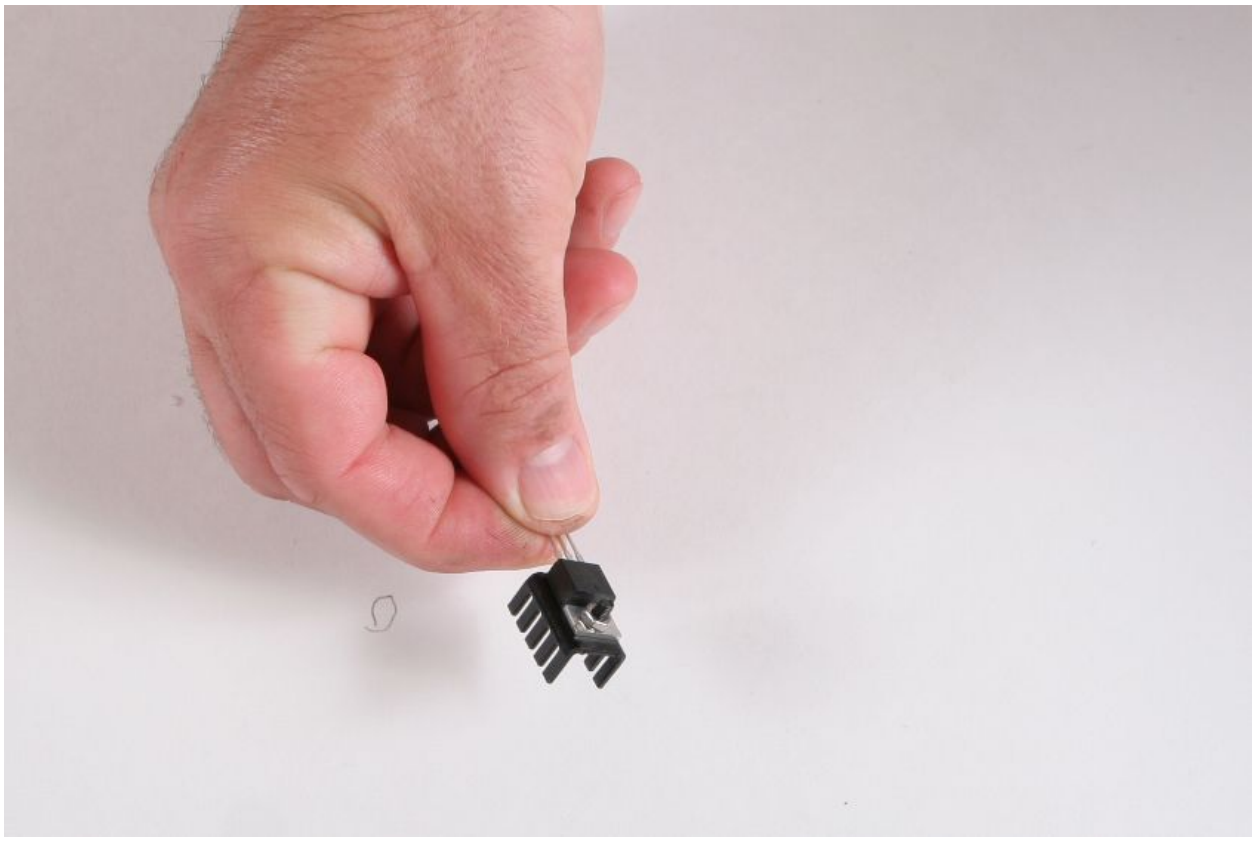
Heat Sink

Parts	Quantity	Image
HEAT SINK TO-220 .375" COMPACT	1	
IC REG LDO 5V 1.5A TO220	1	

Screw the heat-sink to the 7805 voltage regulator.



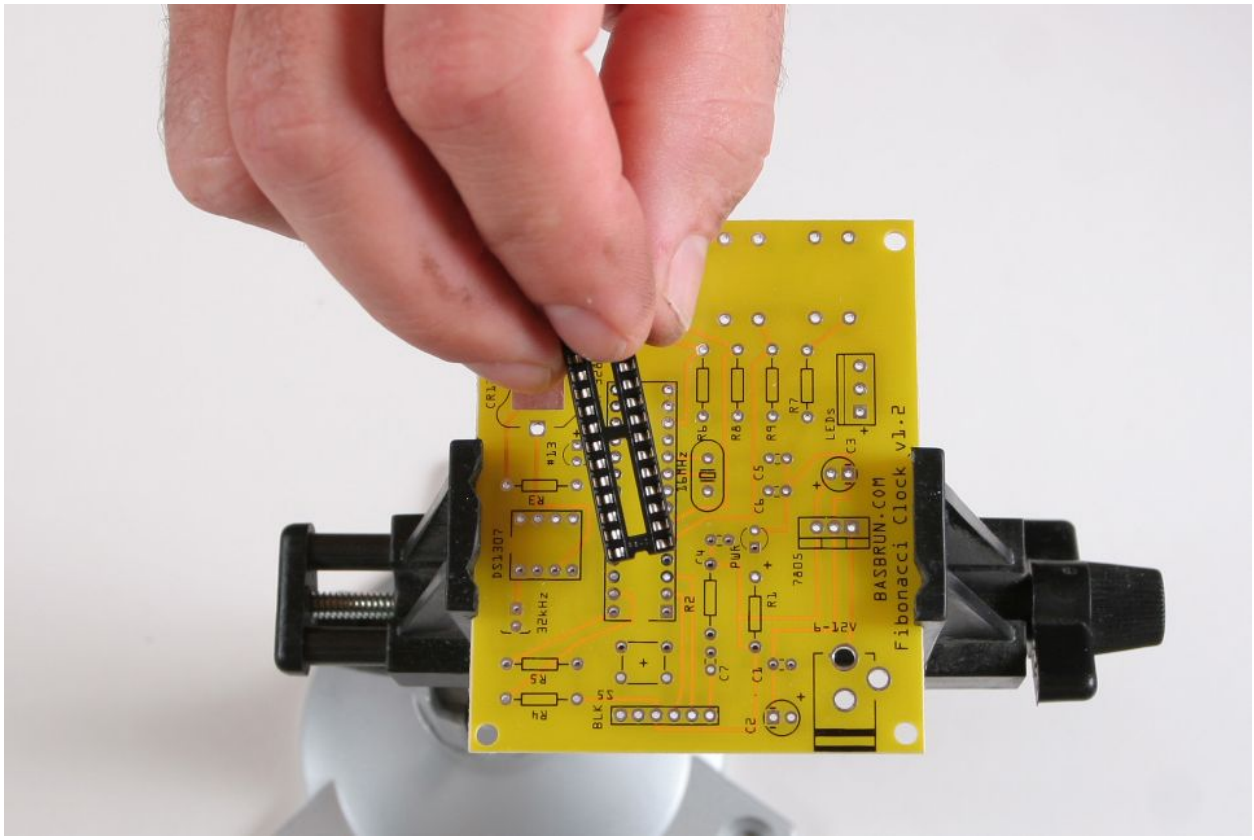


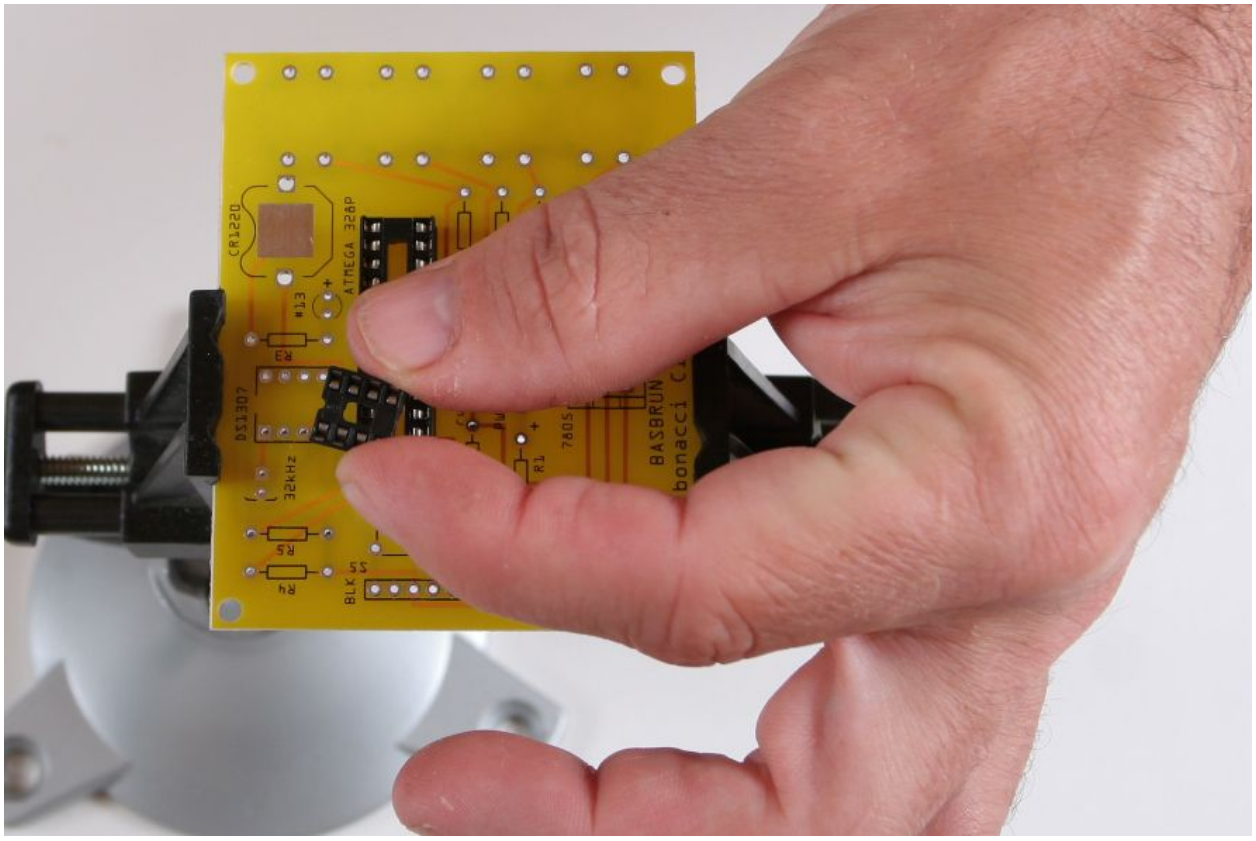


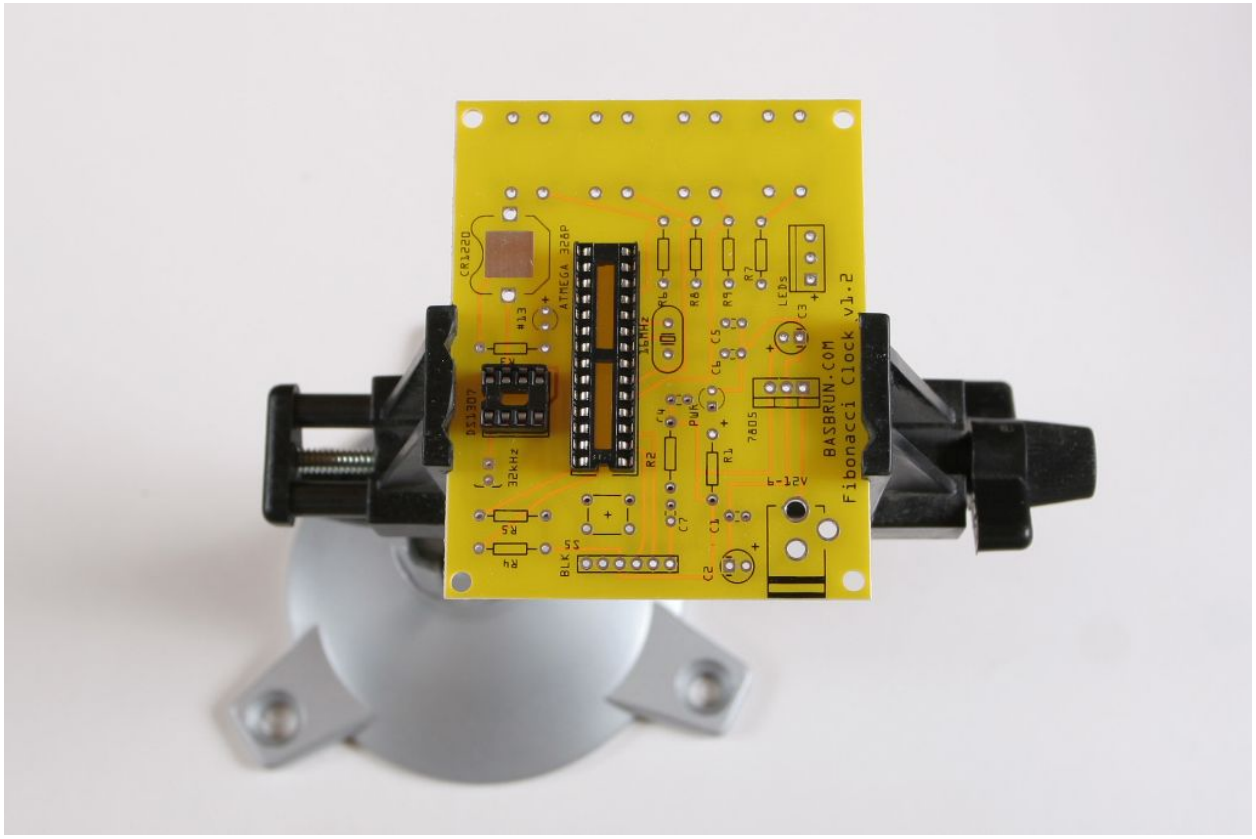
DIP Sockets

Parts	Quantity	Image
CONN IC DIP SOCKET 28POS TIN	1	
CONN IC DIP SOCKET 8POS TIN	1	

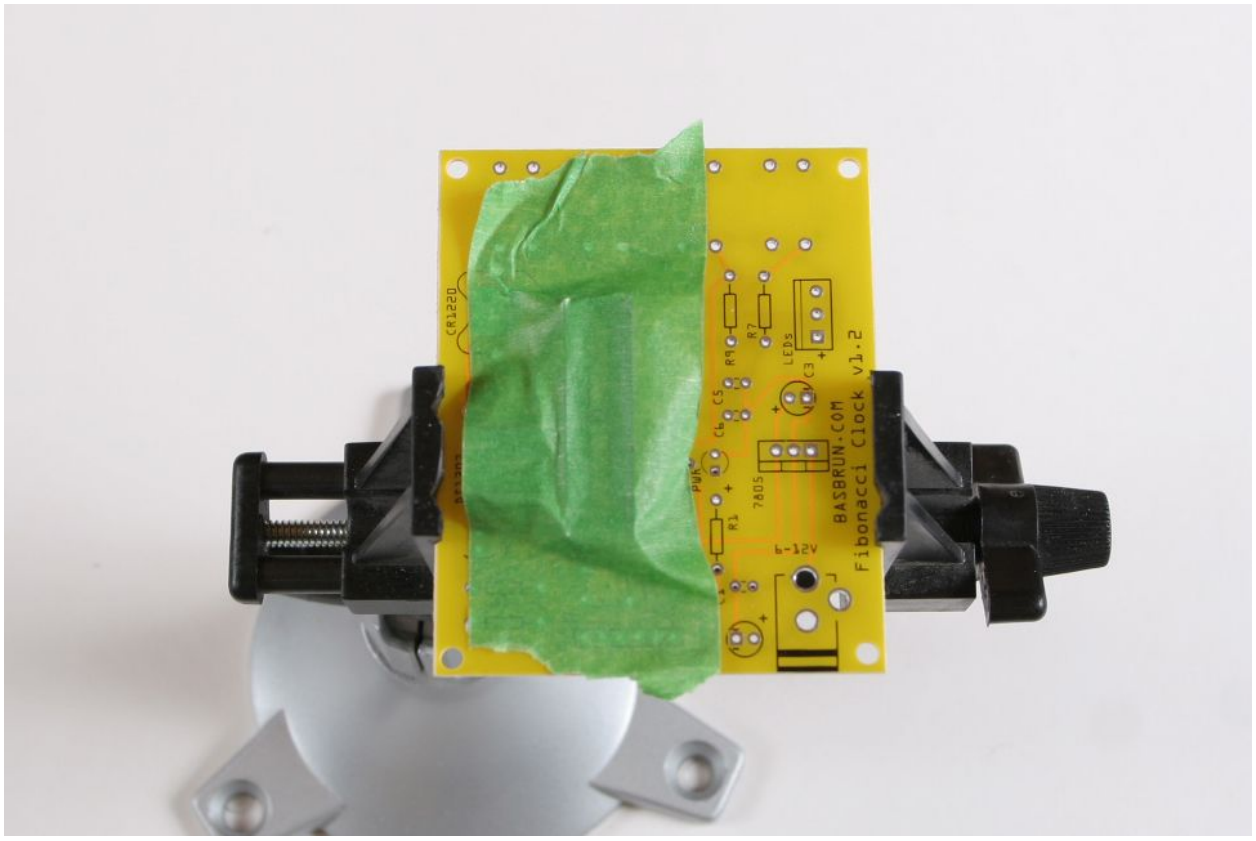
Place the two DIP sockets in their locations.. The dent at the end of the DIP sockets should align with the dent on the circuit board drawings.



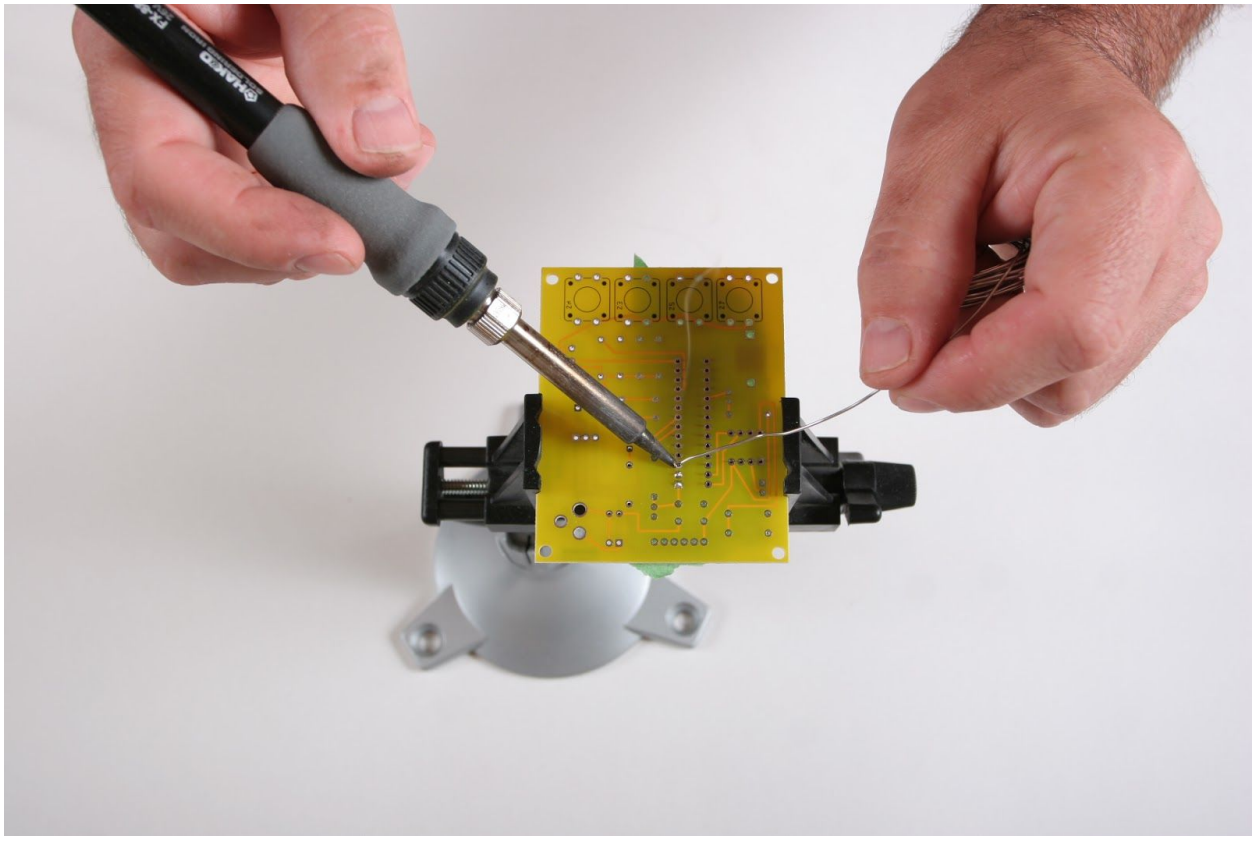

CR1220
ATMEGA 328P
DS1307
32KHz
BLK
7805
BASBRUN
bonacci

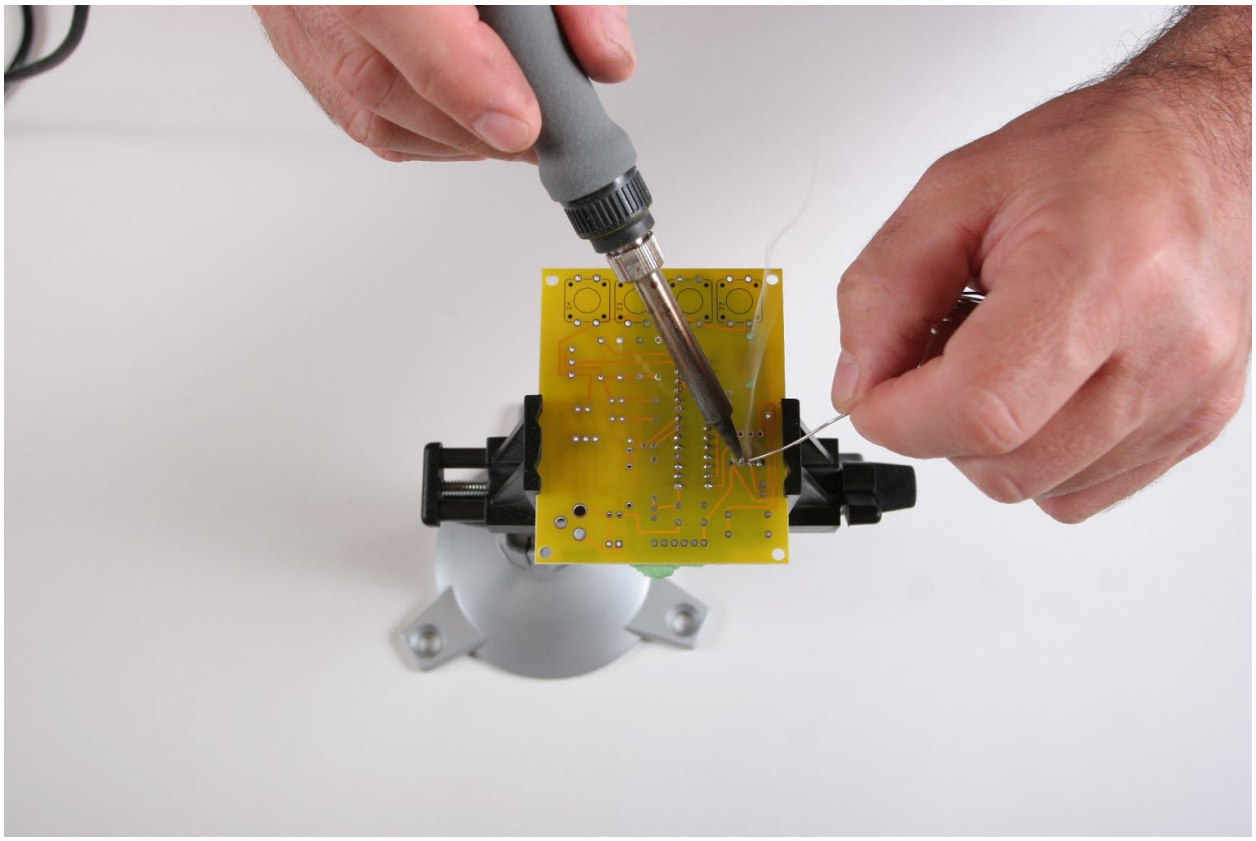

CR1220
ATMEGA 328P
DS1307
32KHz
BLK
7805
LEDs
BASBRUN.COM
Fibonacci Clock v1.2
5-12V

Secure with tape if necessary.



Solder

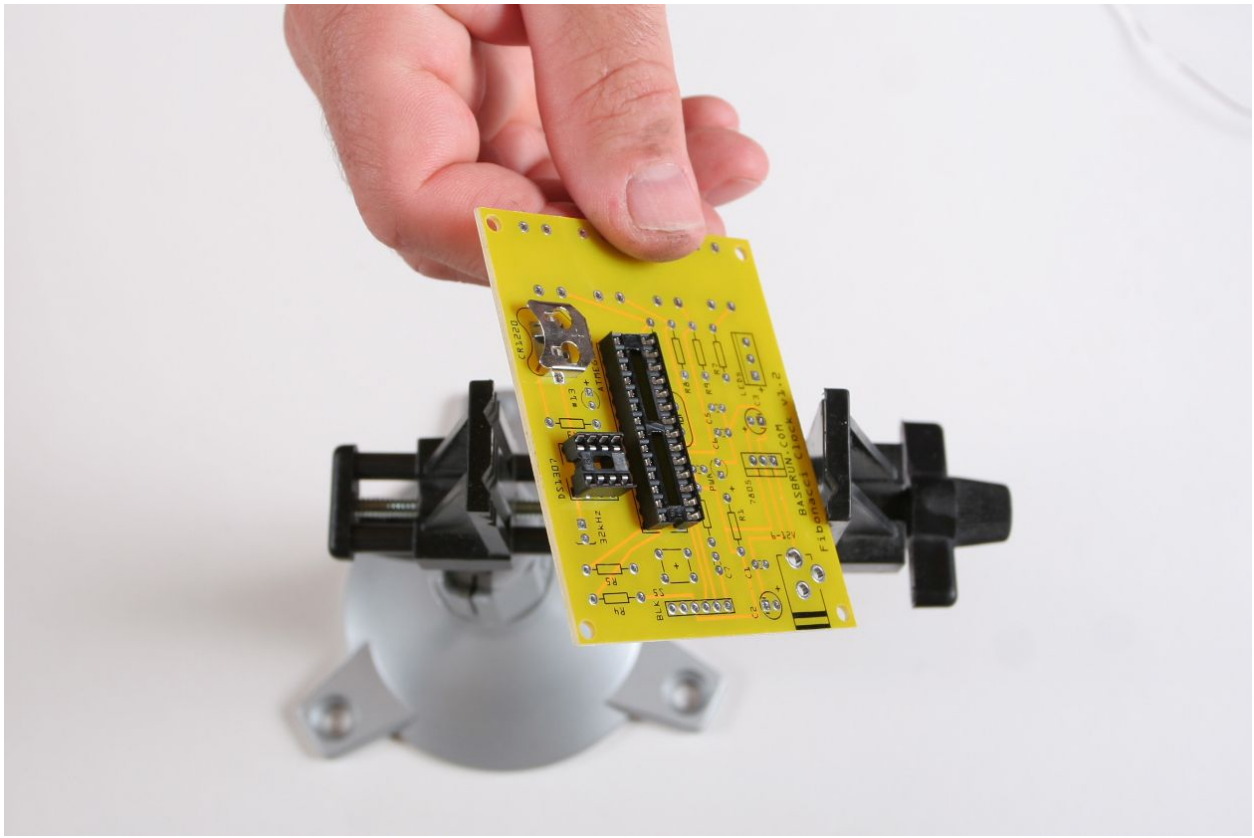




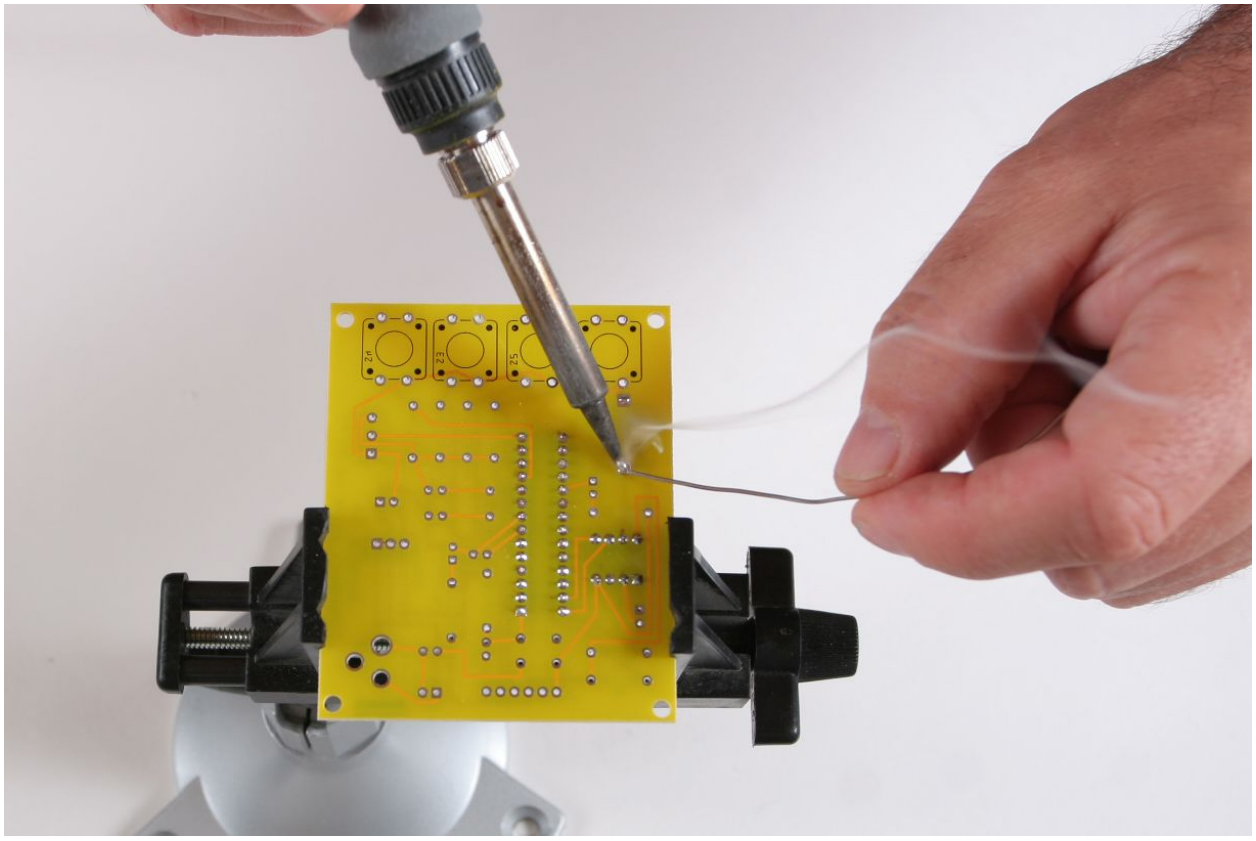
Coin Battery Retainer

Parts	Quantity	Image
RETAINER COIN 12MM W/PC PINS	1	

Snap the coin battery retainer into place. Make sure the opening if facing the edge of the board.



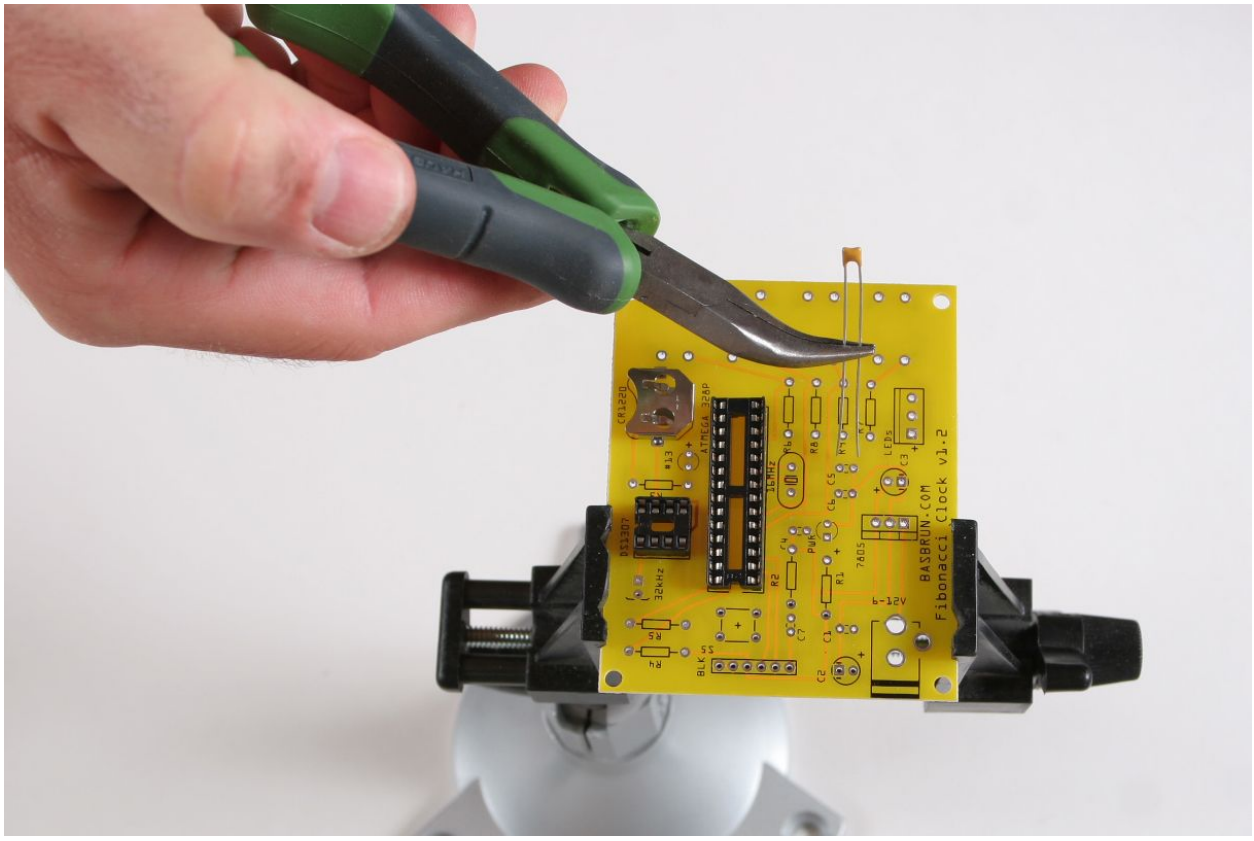
Solder



Ceramic Capacitors

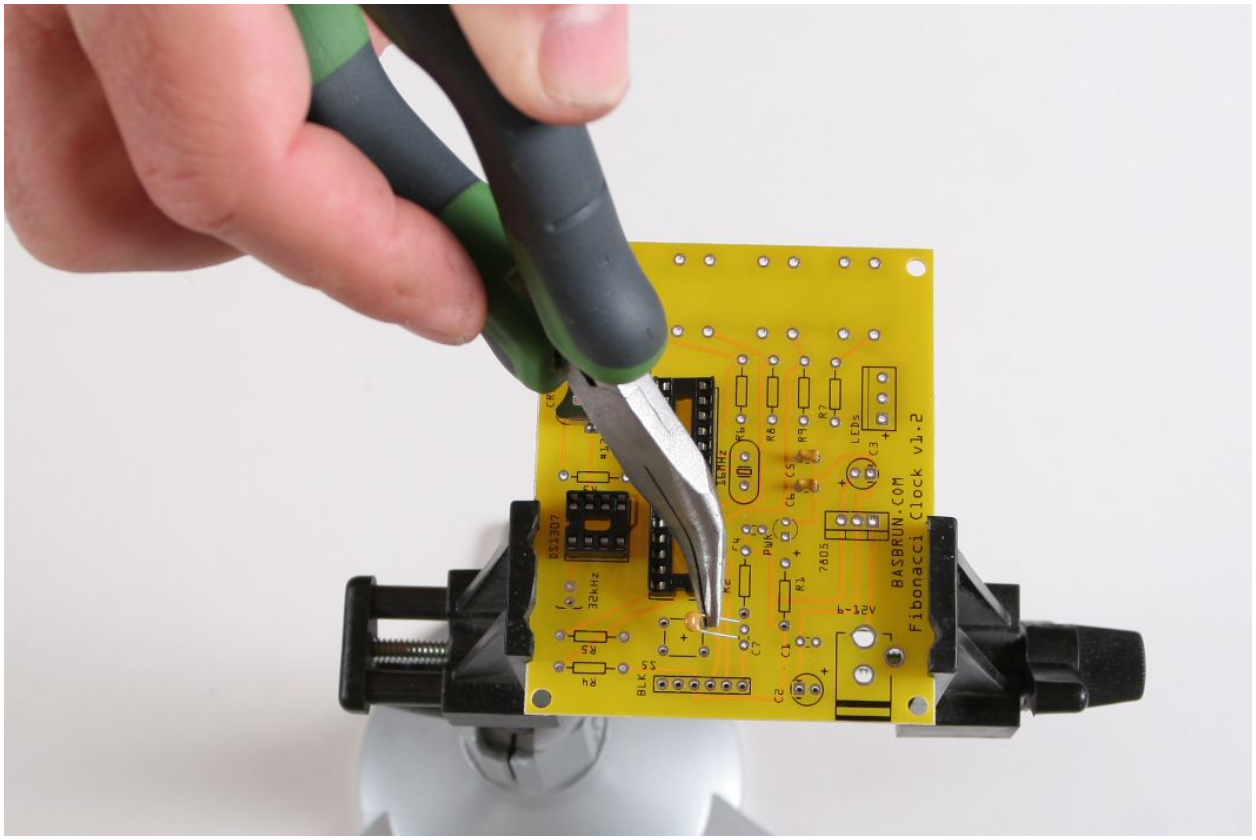
Parts	Quantity	Image
CAP CER 22PF 200V 10% RADIAL ***May be short yellow in your kit***	2	
CAP CER 0.1UF 50V 10% RADIAL ***May be blue in your kit***	3	

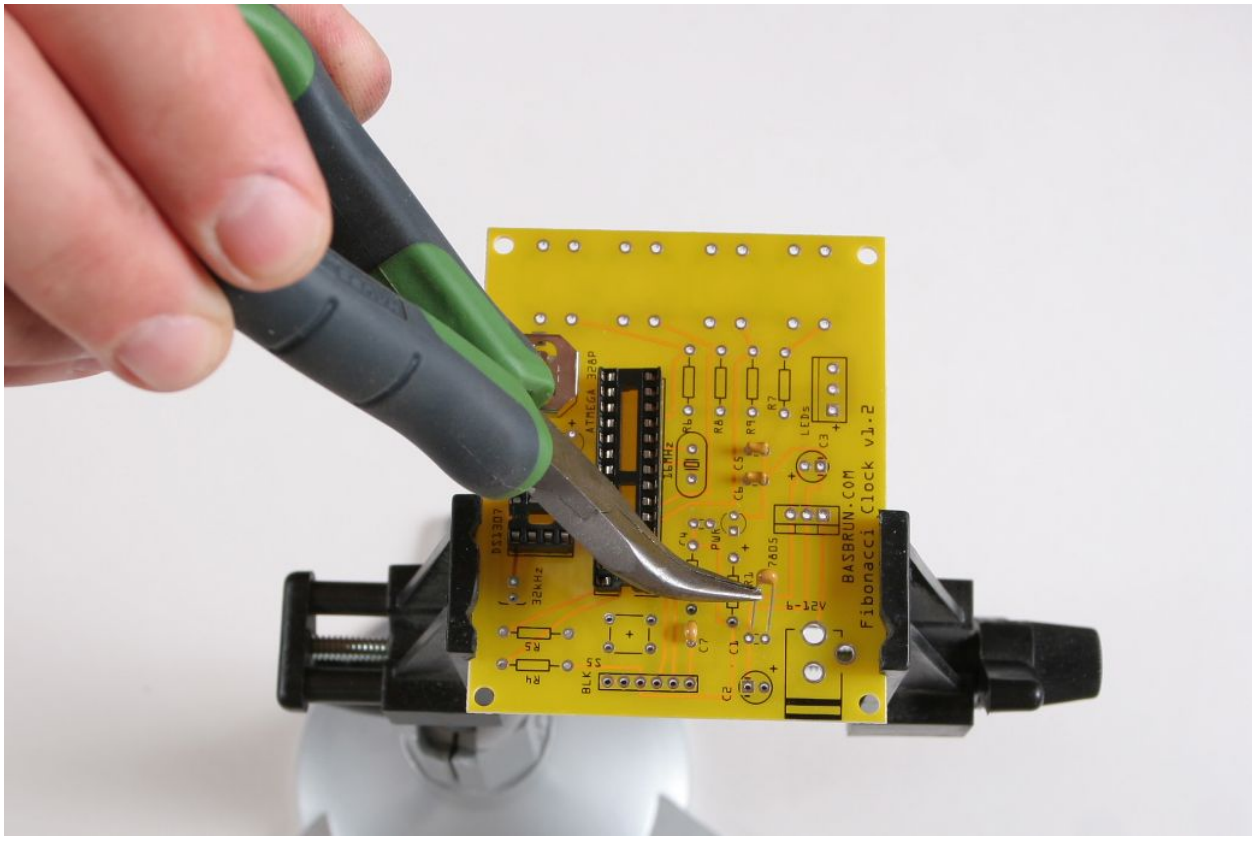
Place the two 22pF capacitors in C5 and C6.
These capacitors are non-polarized and can be soldered in either way.

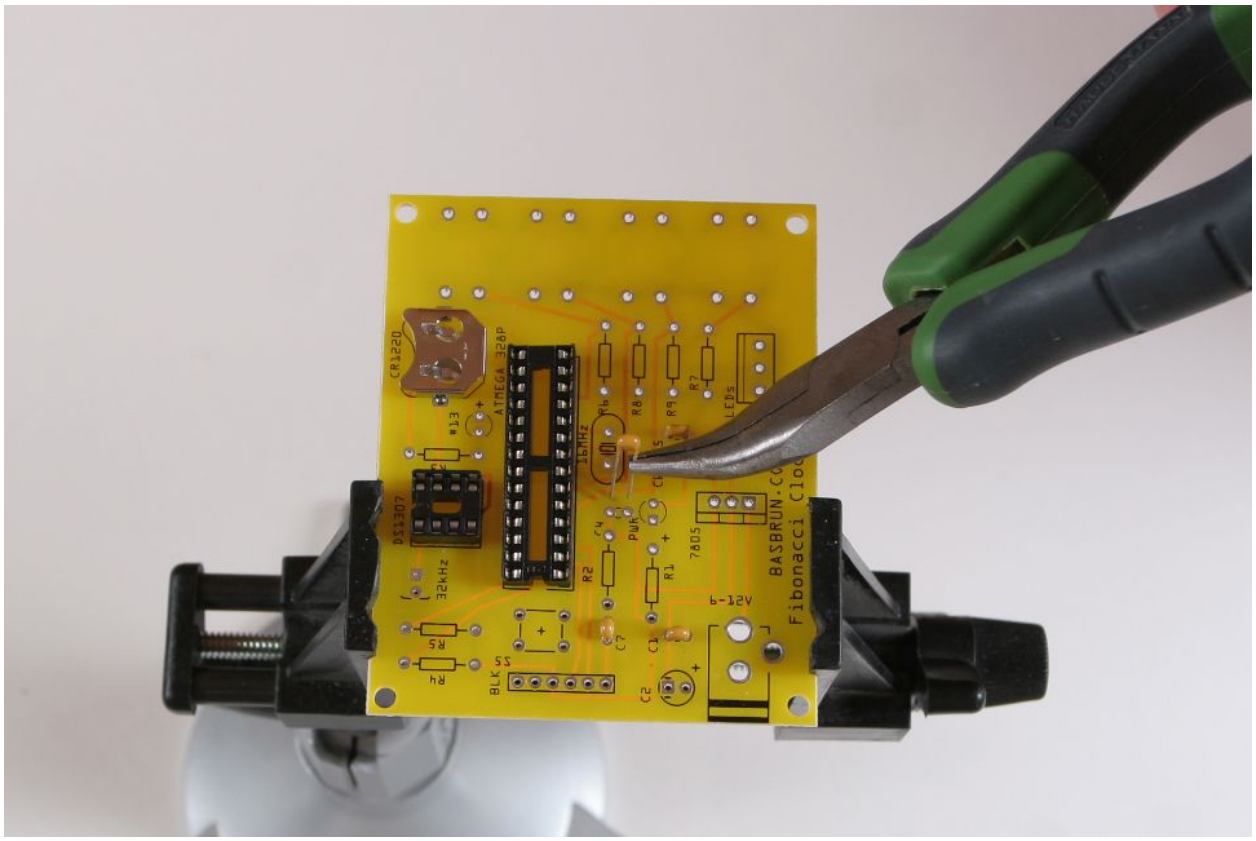

BASBRUN.COM
Fibonacci Clock v1.2

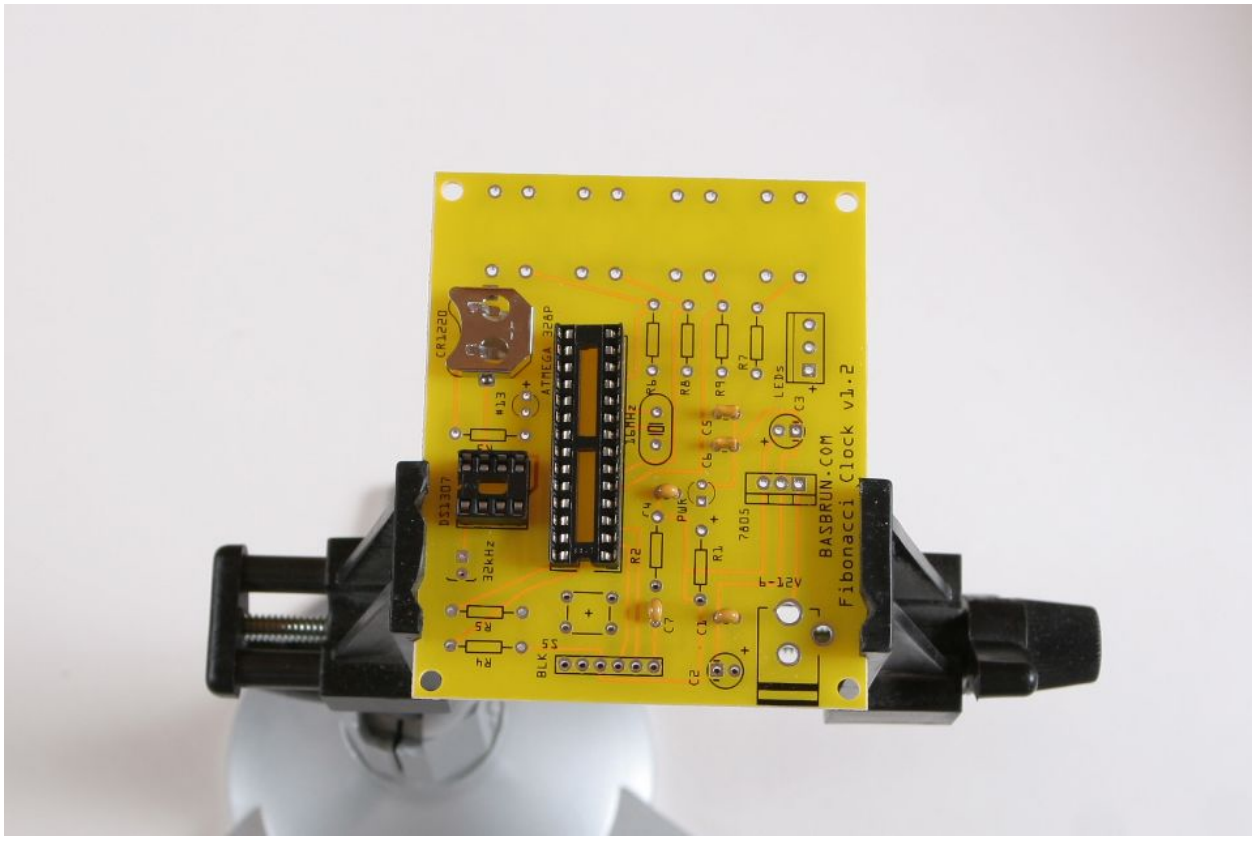
BASBRUN.COM
Fibonacci Clock v1.2

Place the three 0.1uF capacitors in C1, C4 and C7.
These capacitors are non-polarized and can be soldered in either way.

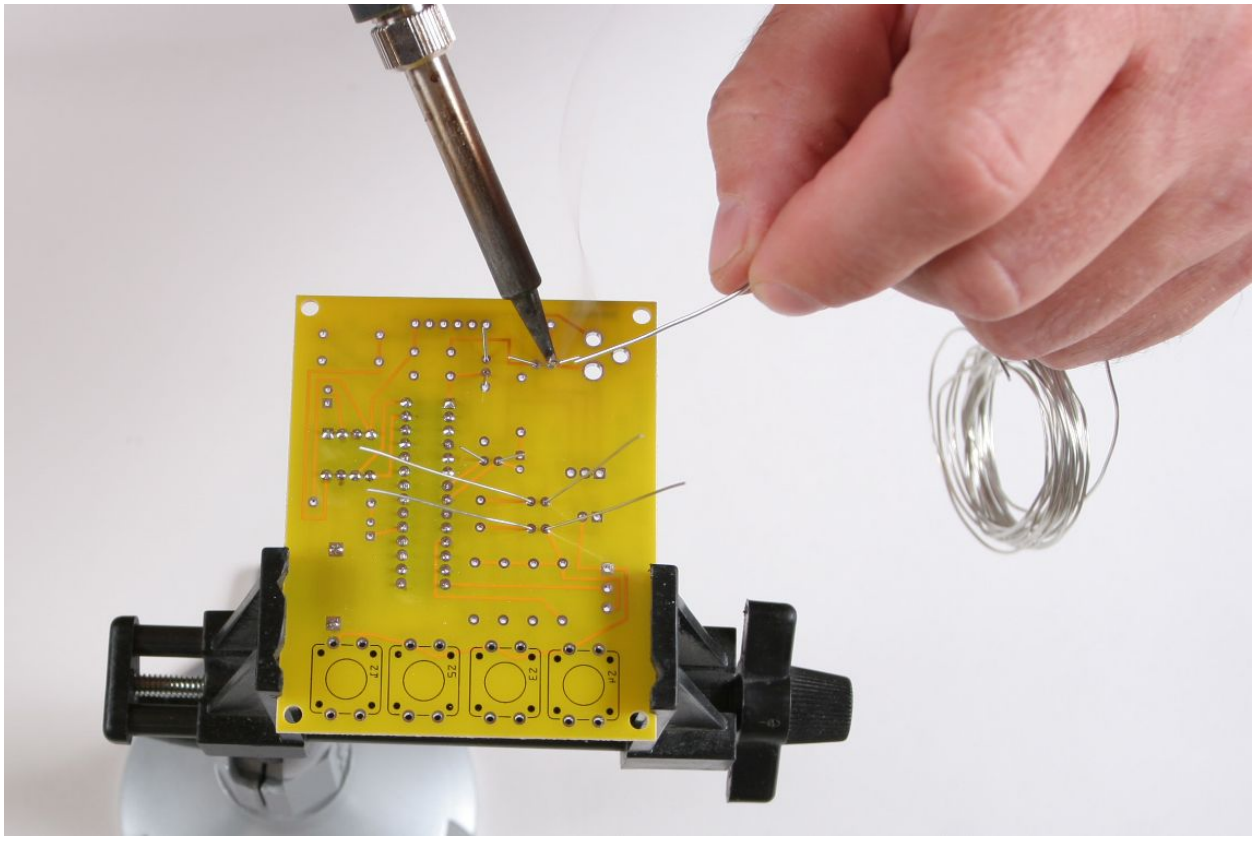



ATMEGA 328P
16MHz
DS1307
32kHz
BLK
7805
6-12V
BASBRUN.COM
Fibonacci Clock v1.2

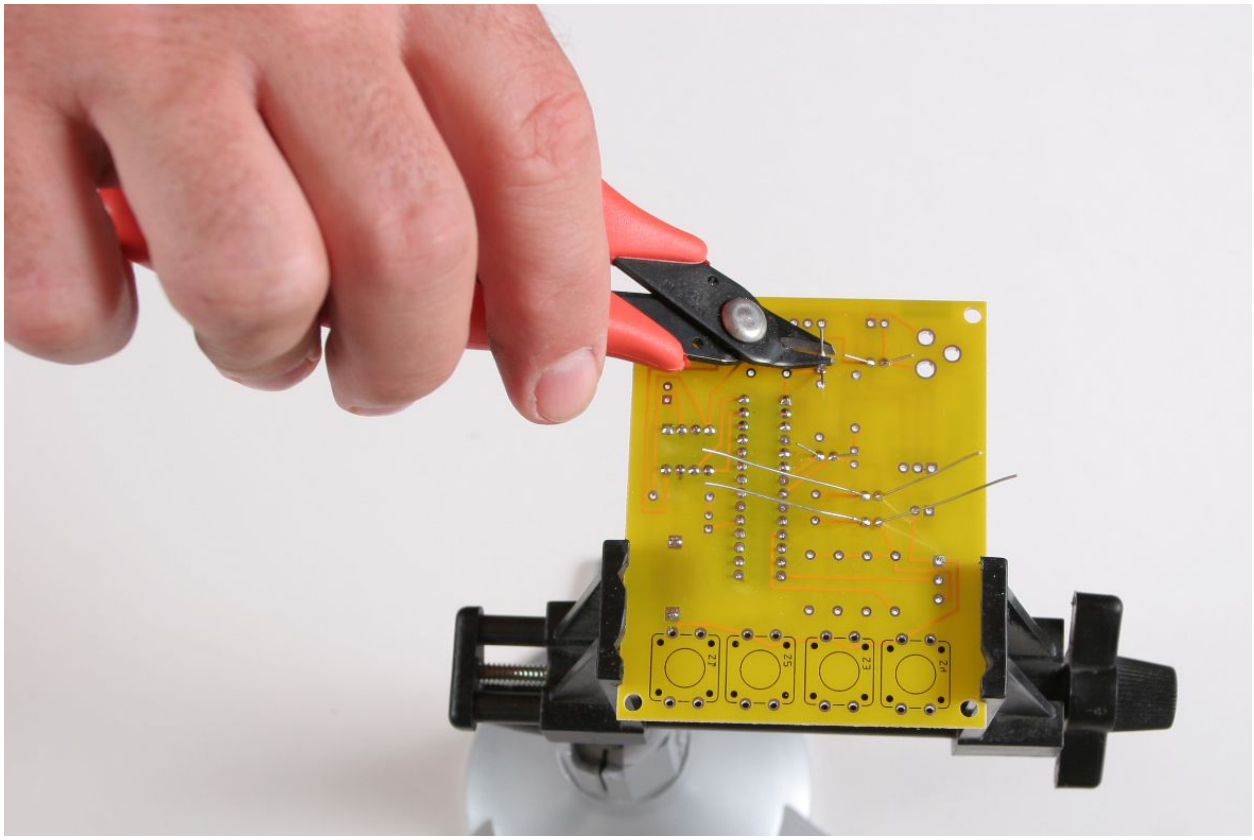

CR1220
ATMEGA 328P
16MHz
DS1307
32kHz
BLK
7805
6-12V


CR1220
ATMEGA 328P
16MHz
DS1307
32kHz
PWR
7805
6-12V
BASBRUN.COM
Fibonacci Clock v1.2

Solder



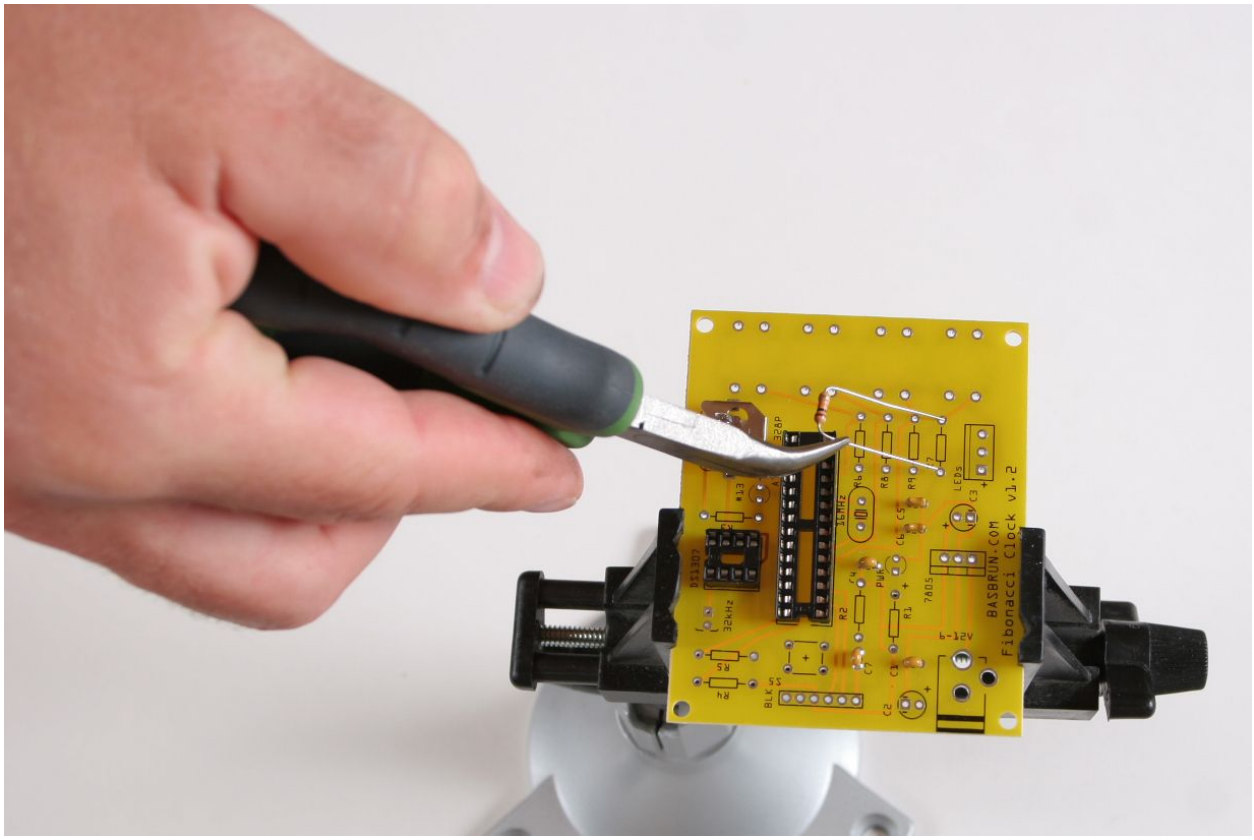
Cut the excess wires.

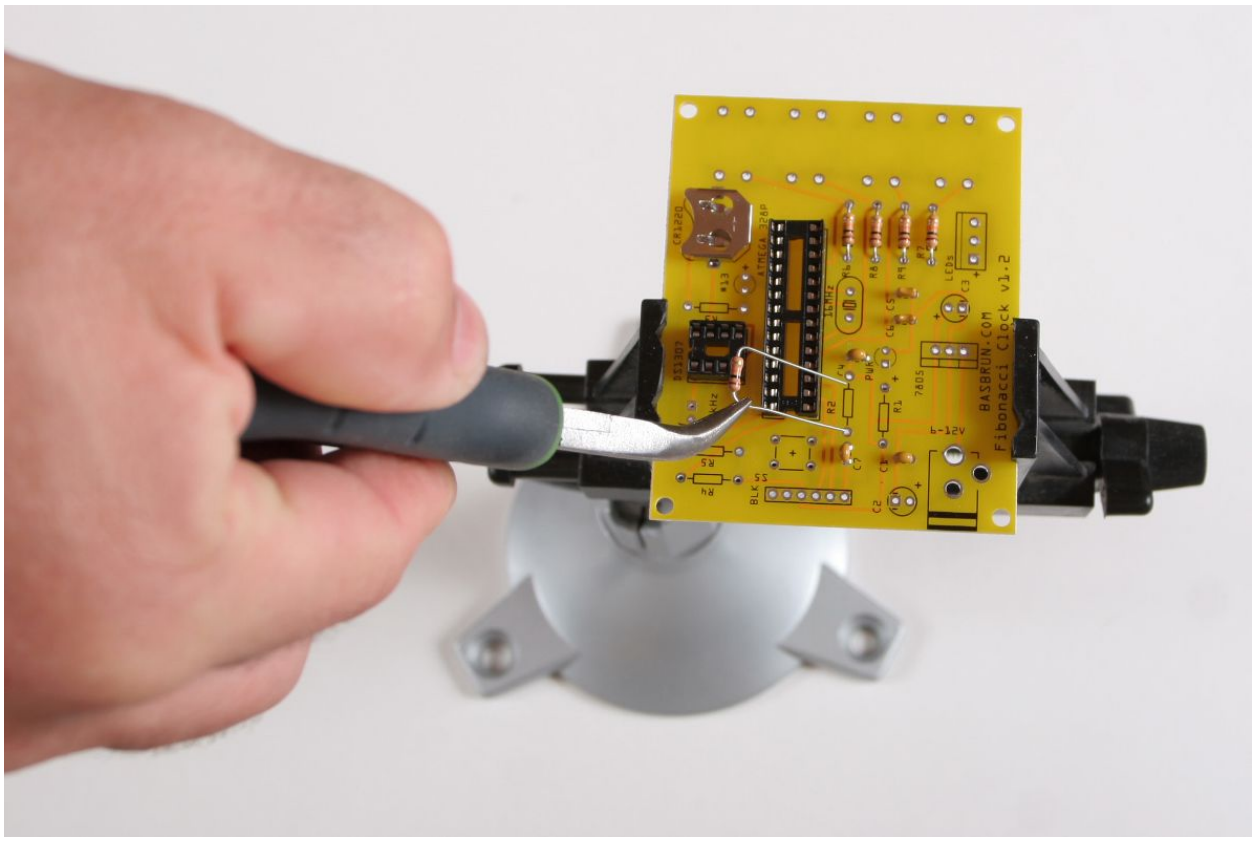


Resistors

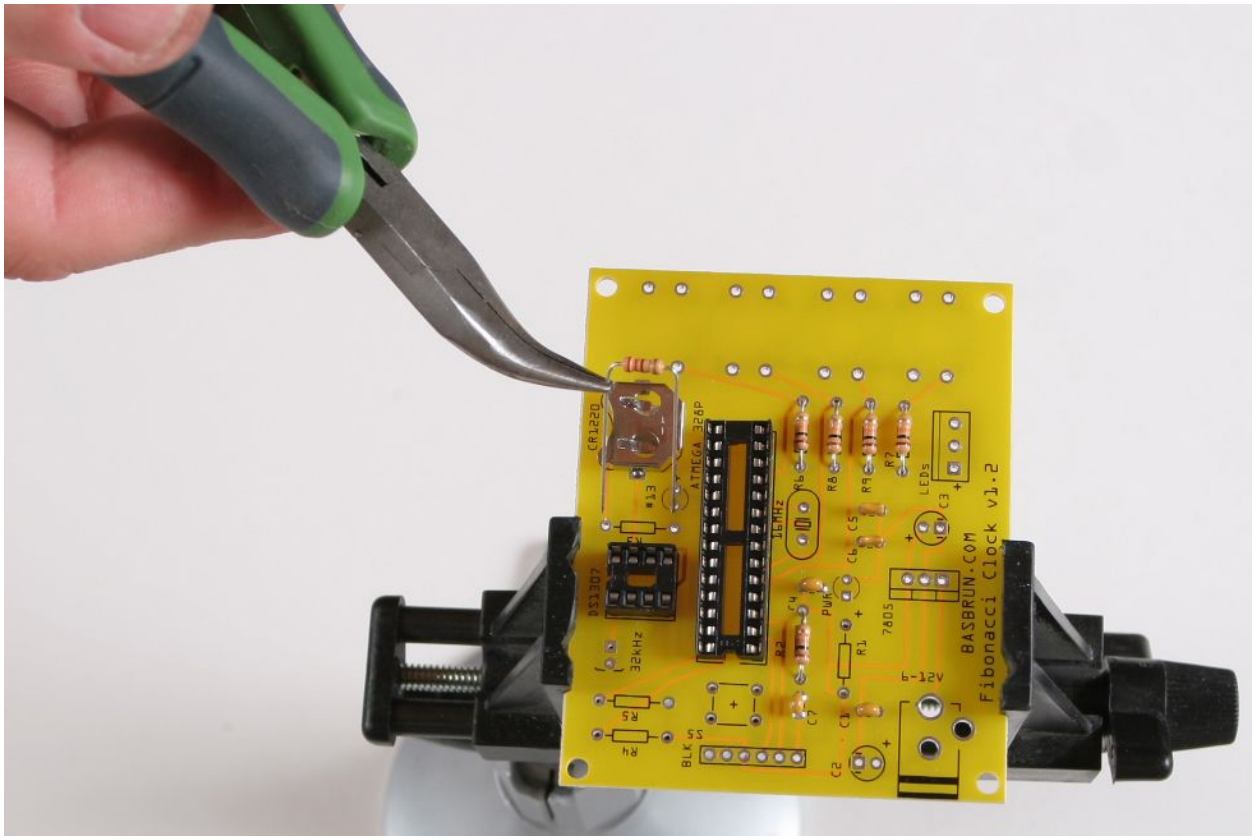
Parts	Quantity	Image
RES 10K OHM 1/4W 5% AXIAL	5	
RES 2.2K OHM 1/4W 5% CARBON FILM	2	
RES 220 OHM 1/4W 5% AXIAL	2	

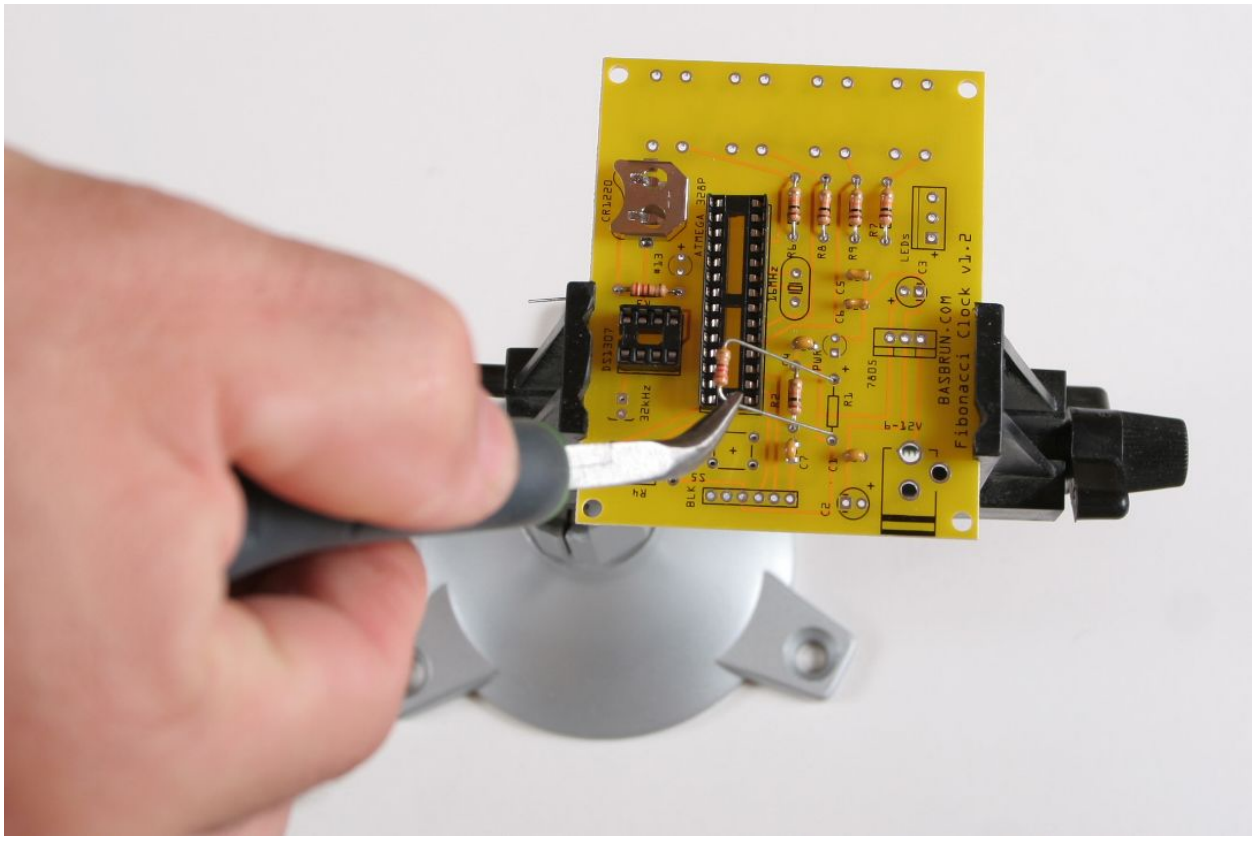
Place the five 10KH resistors in R2, R6, R7, R8 and R9.
Resistors are non-polarized and can be soldered in either way.


BASBRUN.COM
Fibonacci Clock v1.2
6-12V
7805
32KHz
BLK

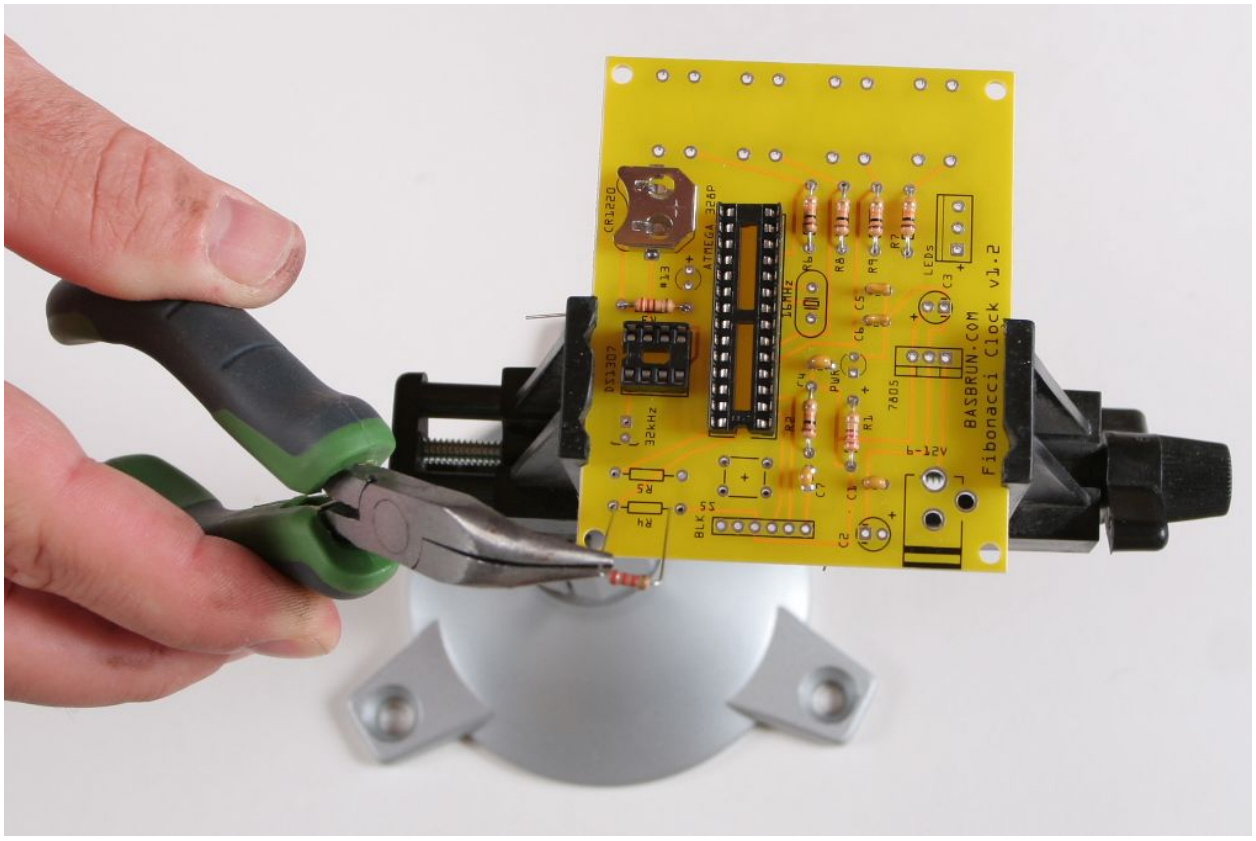

BASBRUN.COM
Fibonacci Clock v1.2
6-12V
7805
32KHz
BLK

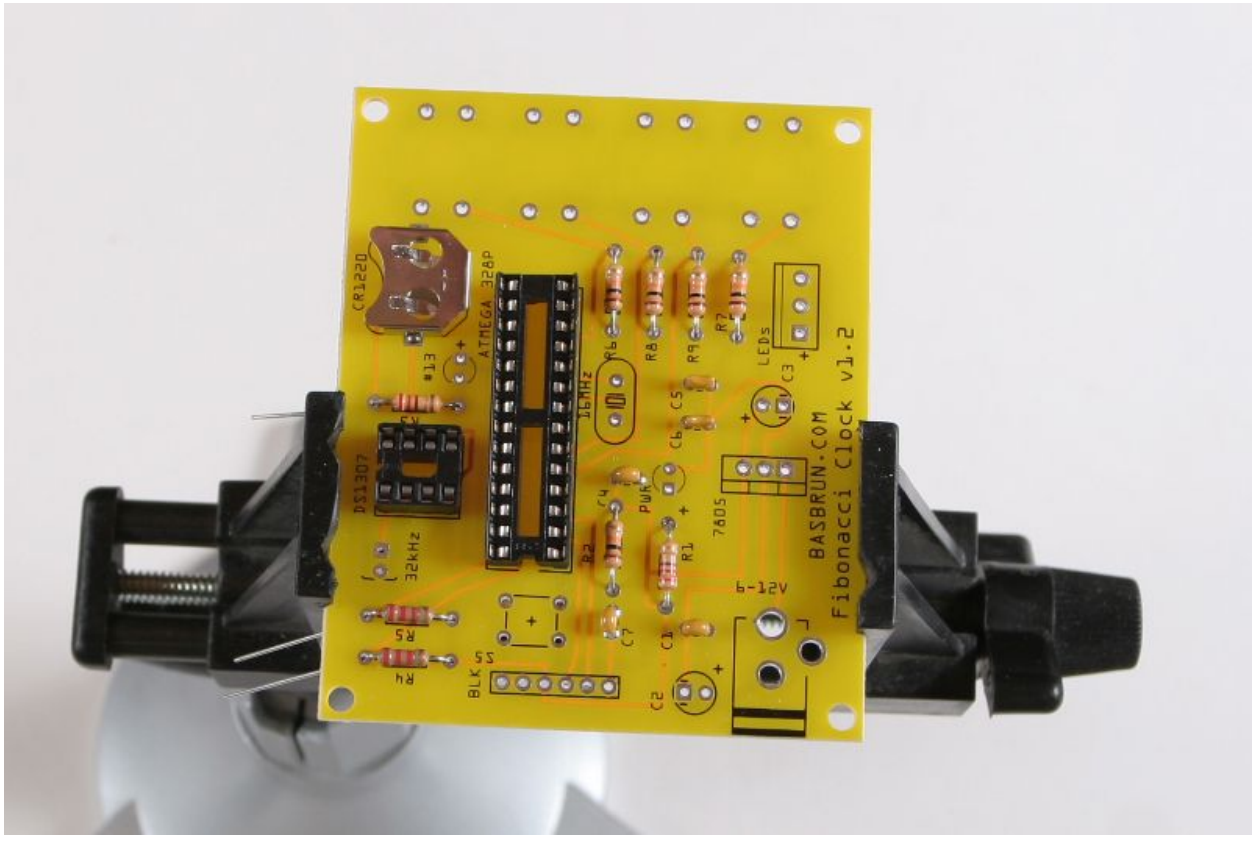
Place the two 220H resistors in R1 and R3.
Resistors are non-polarized and can be soldered in either way



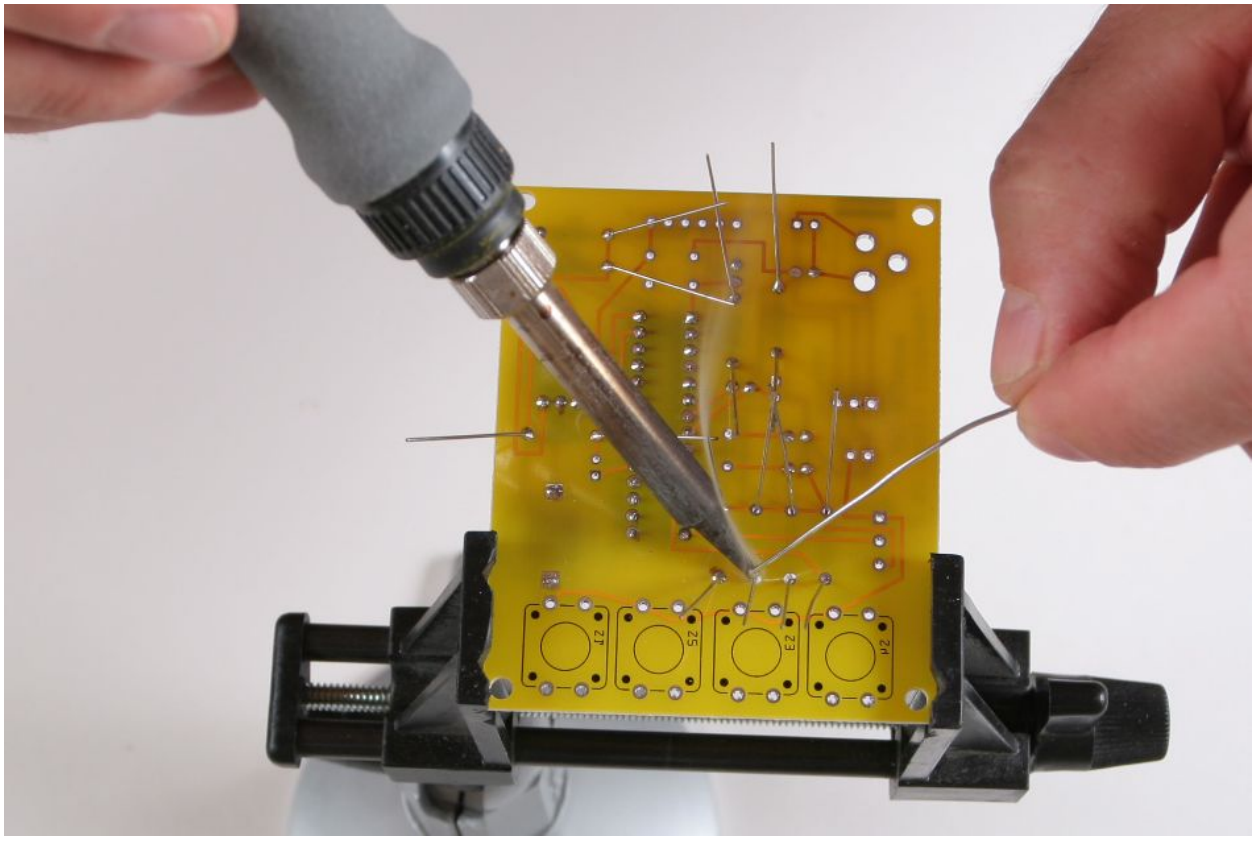


Place the two 2.2KH resistors in R4 and R5.
Resistors are non-polarized and can be soldered in either way

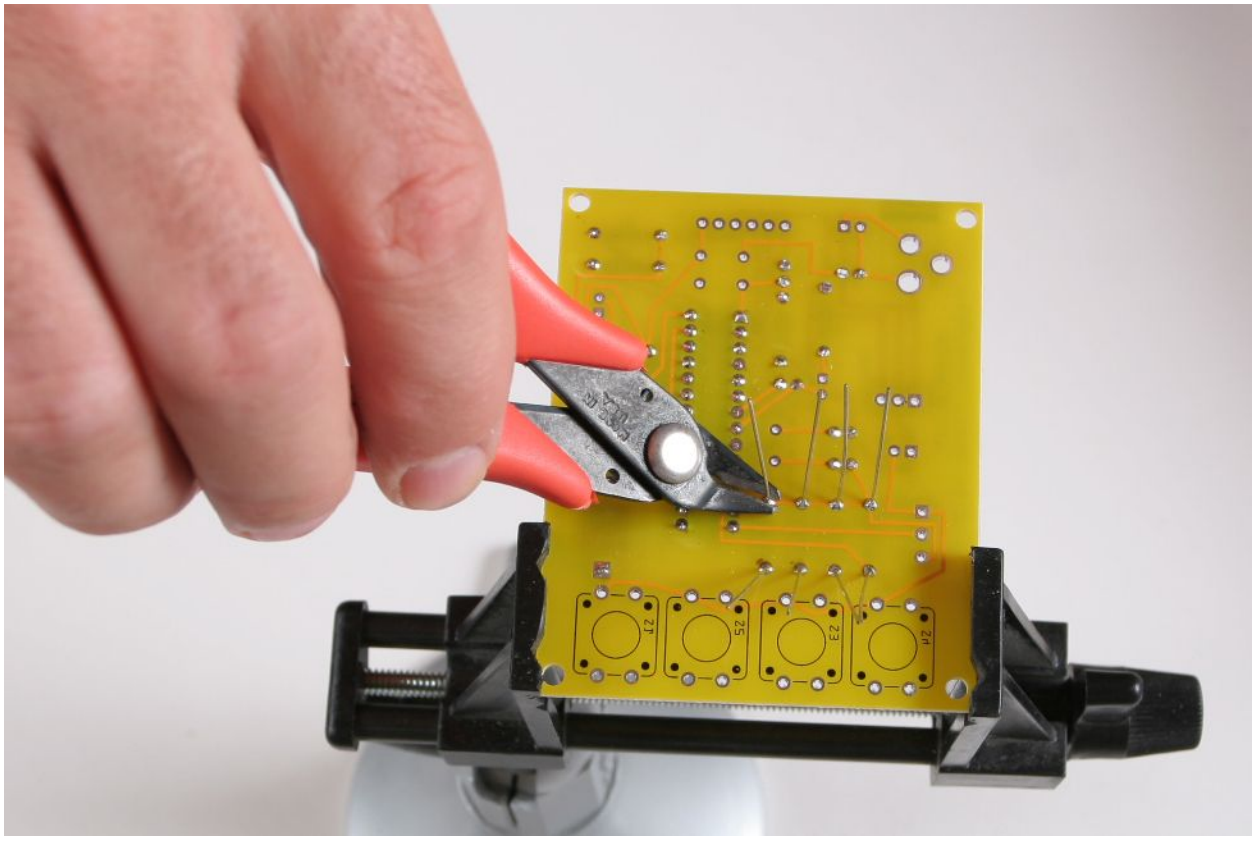

CR1220
ATMEGA 328P
16MHz
DS1307
32kHz
LEDs
7805
6-12V
BASBRUN.COM
Fibonacci Clock v1.2


CR1220
ATMEGA 328P
16MHz
DS1307
32kHz
LEDs
7805
6-12V
BASBRUN.COM
Fibonacci Clock v1.2

Solder



Cut the excess wires.

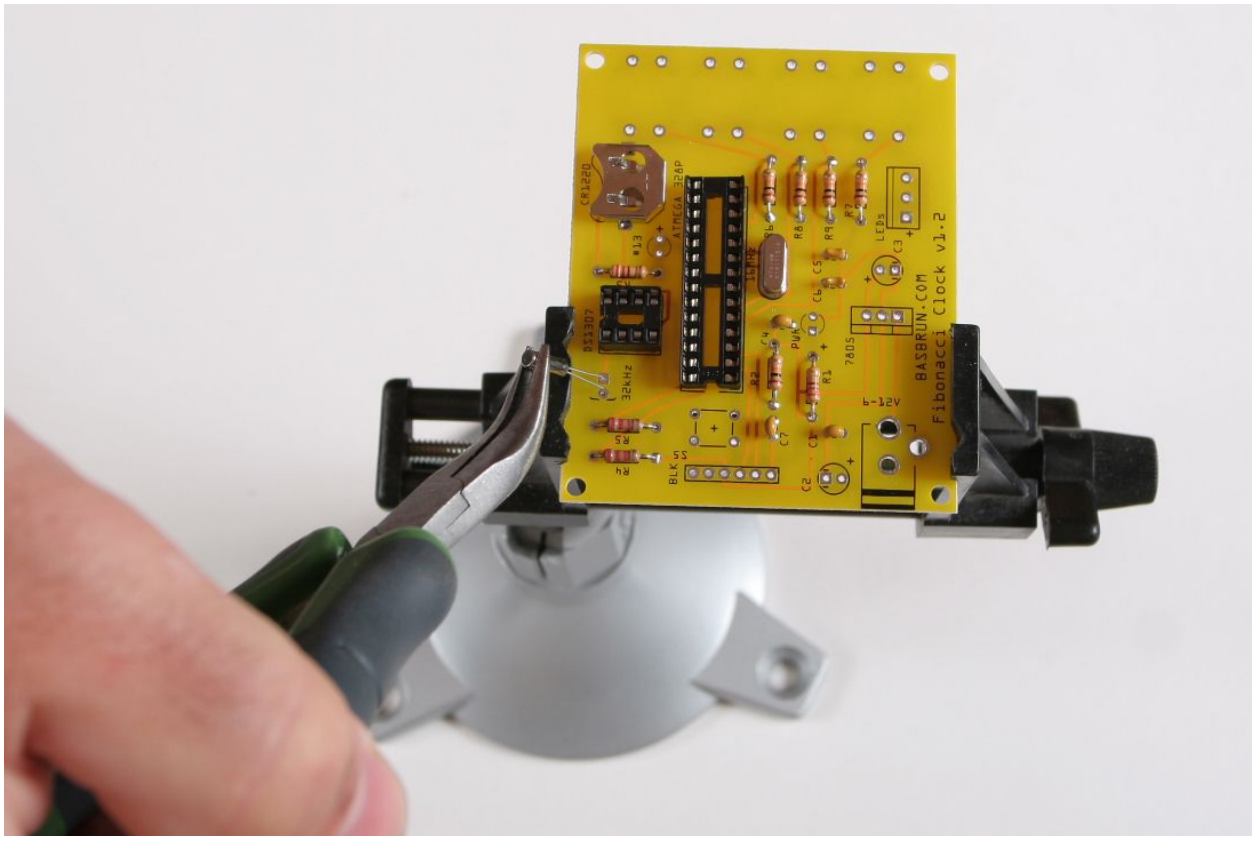


Crystals

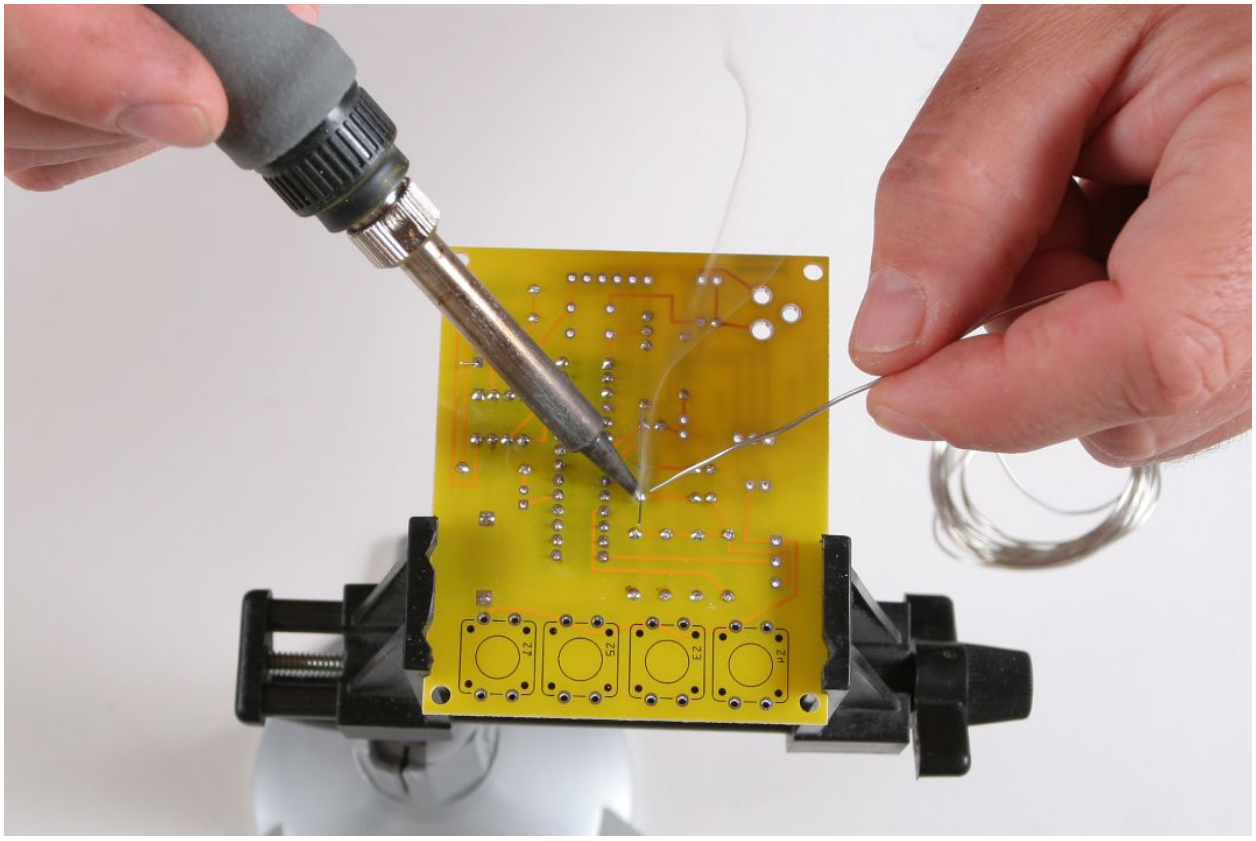
Parts	Quantity	Image
CRYSTAL 16.0000MHZ 18PF T/H	1	
CRYSTAL 32.7680KHZ 12.5PF T/H	1	

Place the 16MHz crystal in it’s location next to the Atmega328P DIP socket.
Crystals are non-polarized and can be placed in either way.

Place the 32KHz crystal in it’s place next to the DS1307 DIP socket.
Crystals are non-polarized and can be placed in either way.
Secure the crystal with tape if necessary.



Solder

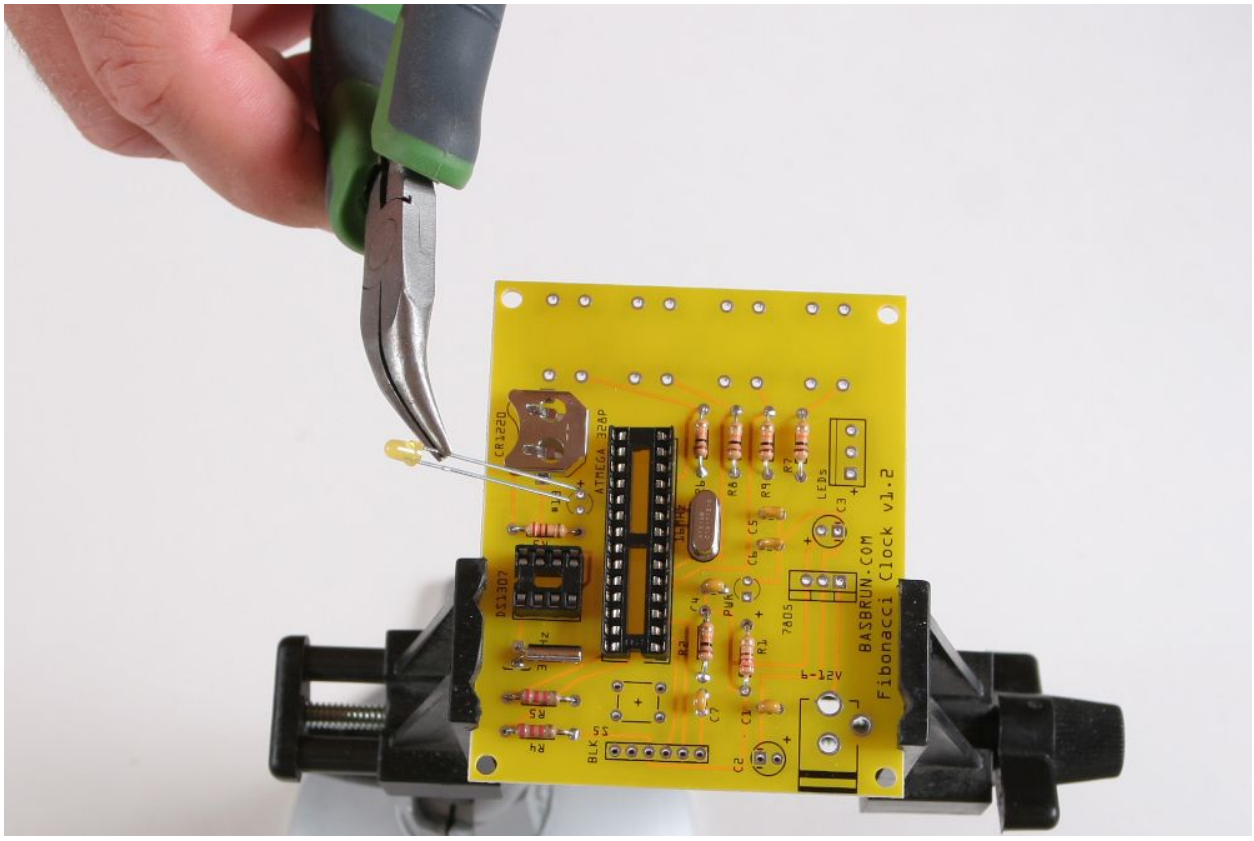


Cut the crystal excess wires.

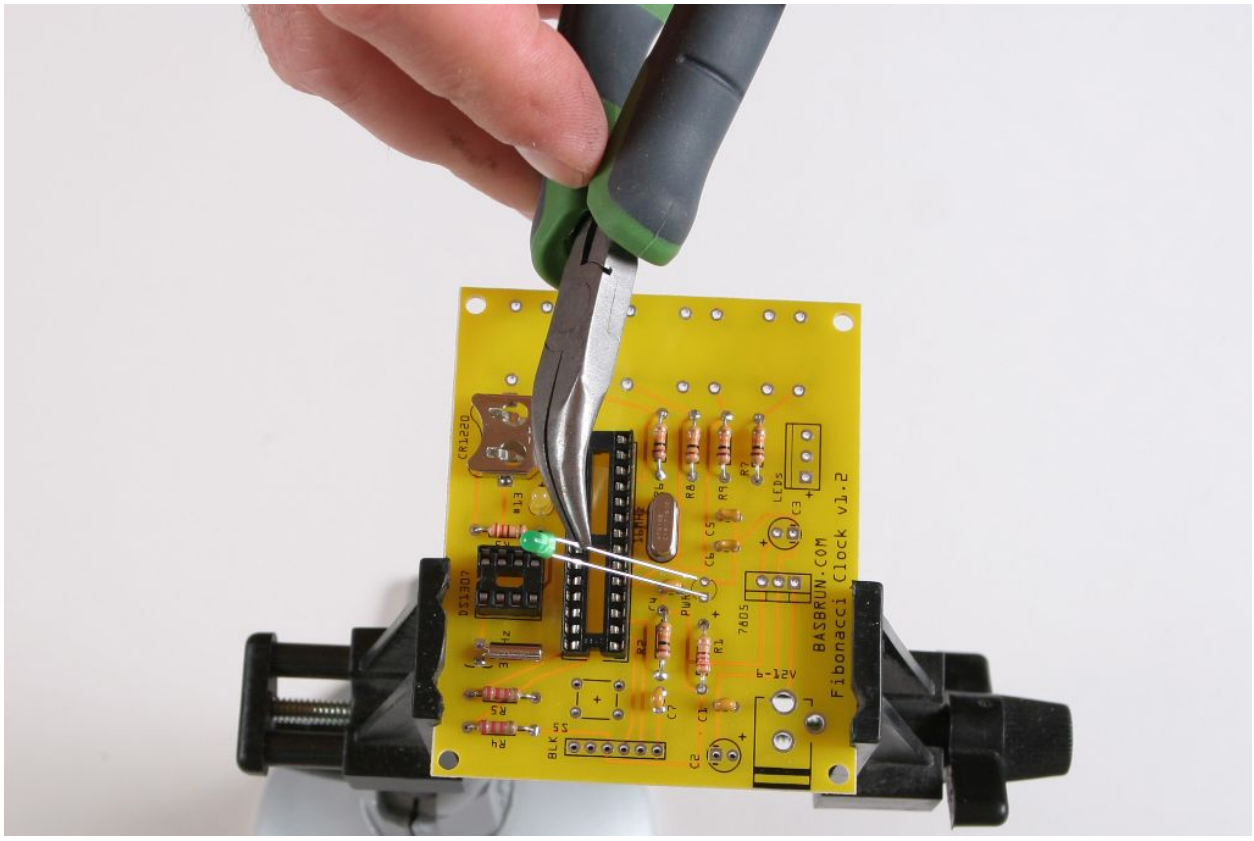
LEDs

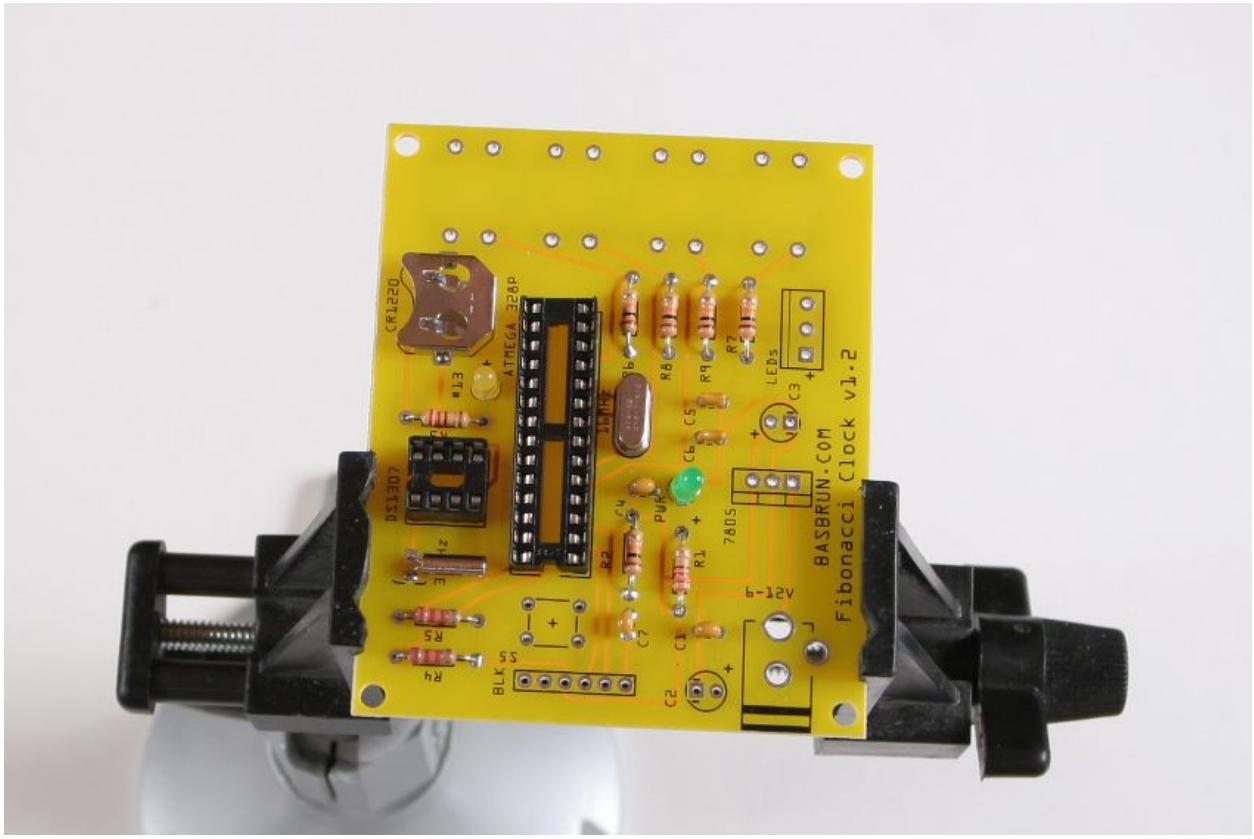
Parts	Quantity	Image
WL-TMRC THT LED ROUND COLOR DIFF	1	
LED GRN 572NM DIFF 3MM RND	1	

Place the yellow LED in its location close to the battery retainer. Make sure the long wire (anode) is in the hole next to the (+) sign.

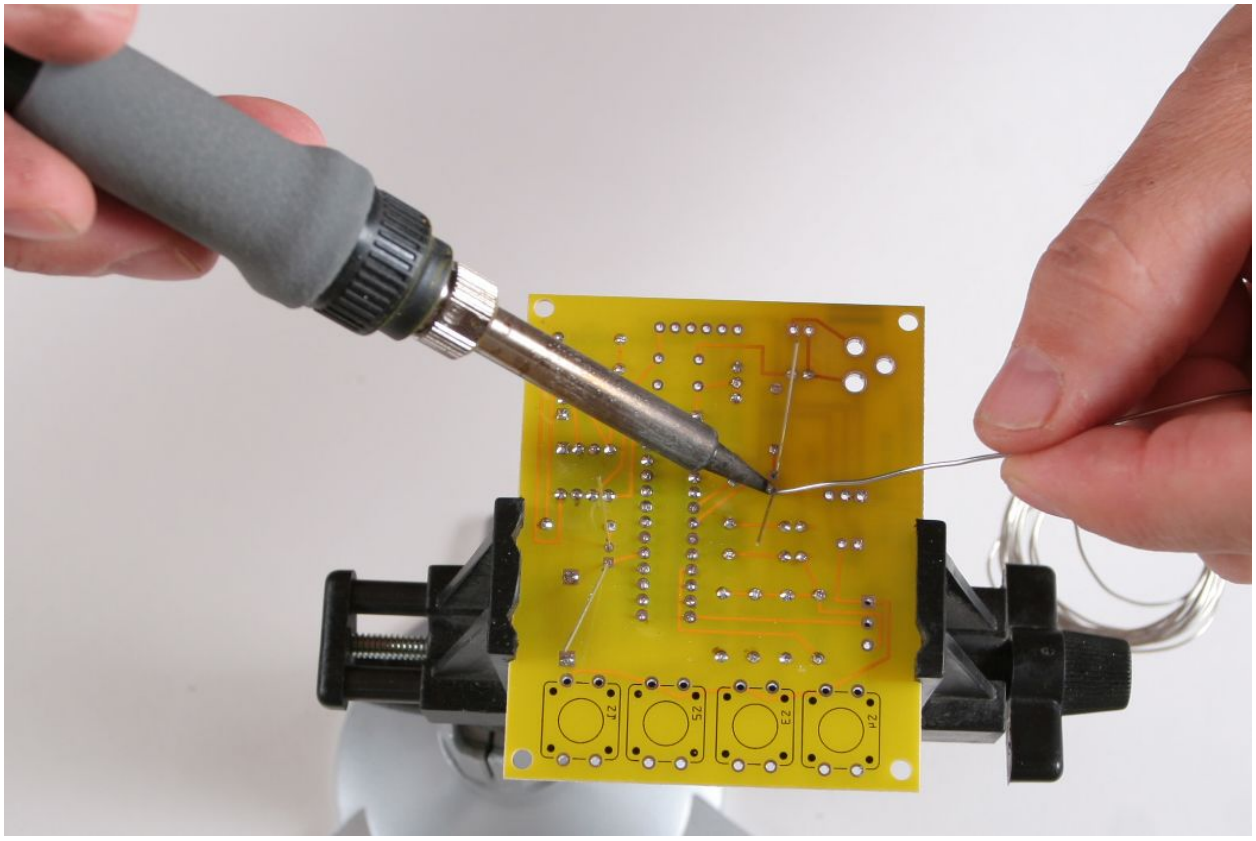


Place the green LED in its location in the center of the board. Make sure the long wire (anode) is in the hole next to the (+) sign.

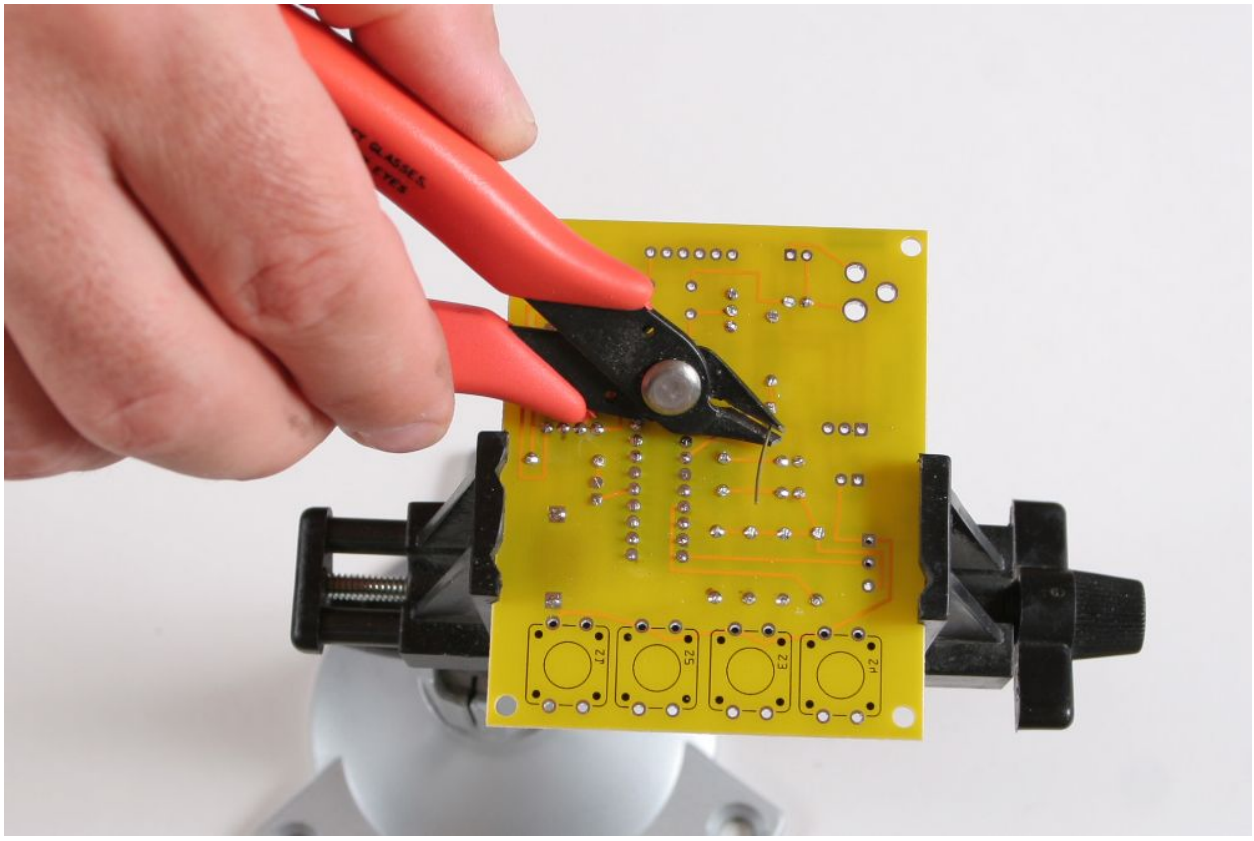

CR1220
DS1307
16MHz
7805
6-12V
BASBRUN.COM
Fibonacci Clock v1.2


CR1220
ATMEGA 328P
DS1307
16MHz
7805
6-12V
BASBRUN.COM
Fibonacci Clock v1.2

Solder



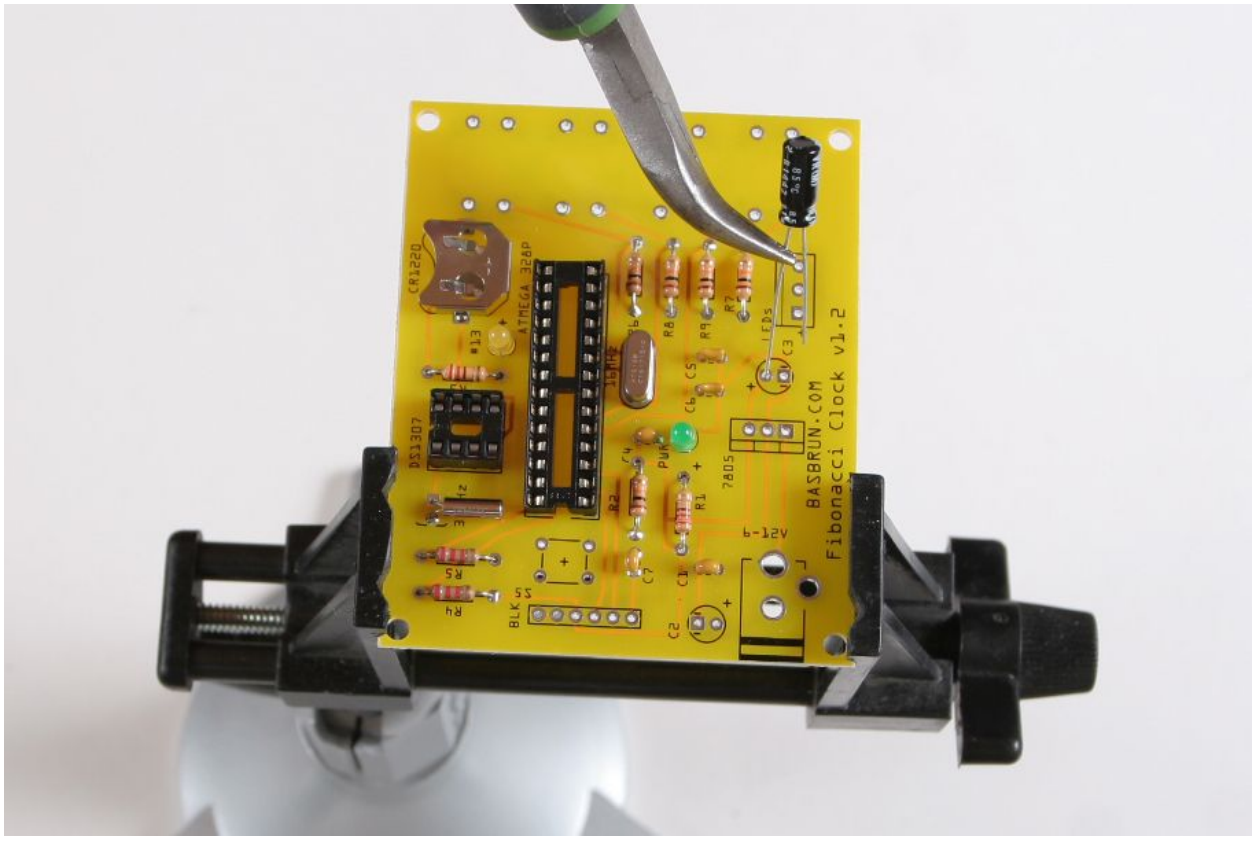
Cut the excess wires.



Electrolytic Capacitors

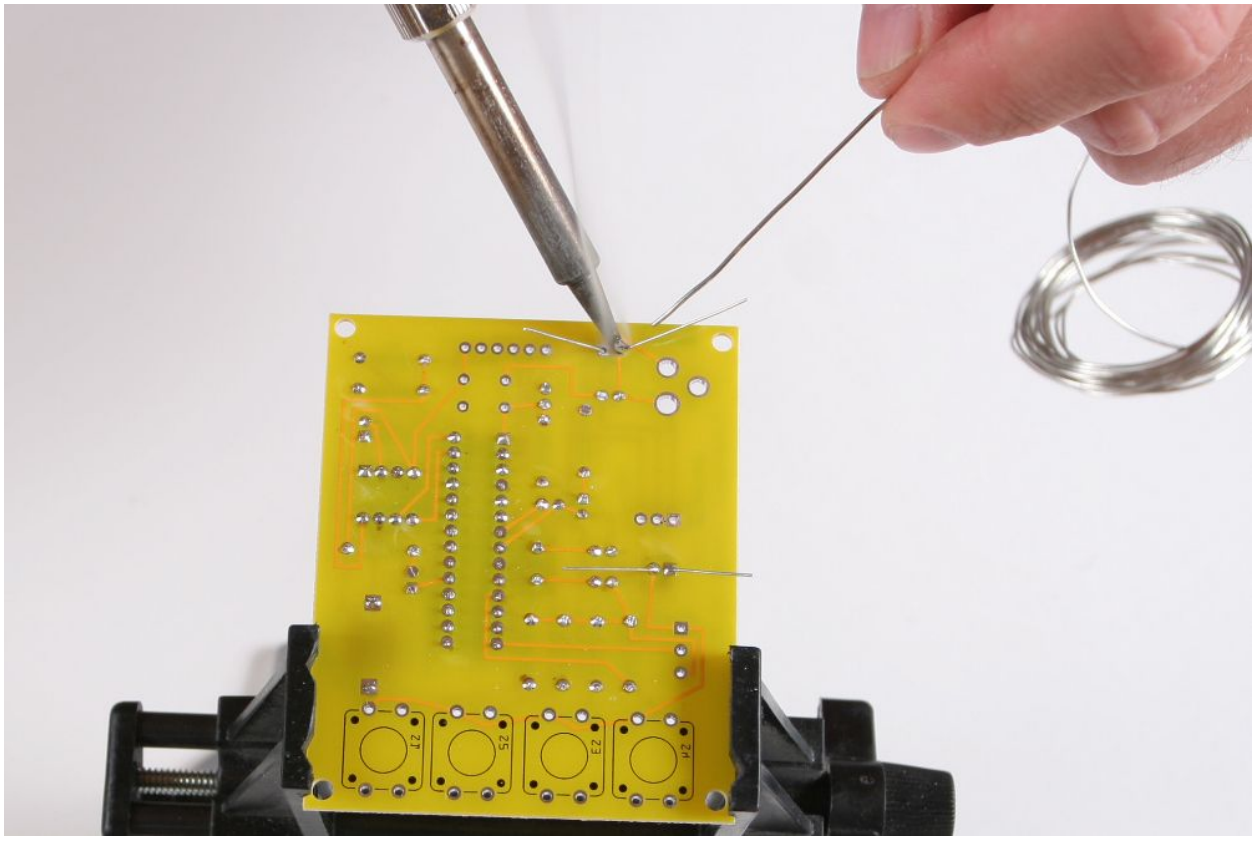
Parts	Quantity	Image
CAP ALUM 100UF 20% 16V RADIAL	1	
CAP ALUM 10UF 20% 25V RADIAL ***May be blue in your kit***	1	

Place the 10uF electrolytic capacitors on C3. Electrolytic capacitors are polarized, make sure to insert the long wire into the round hole marked with a (+) sign.

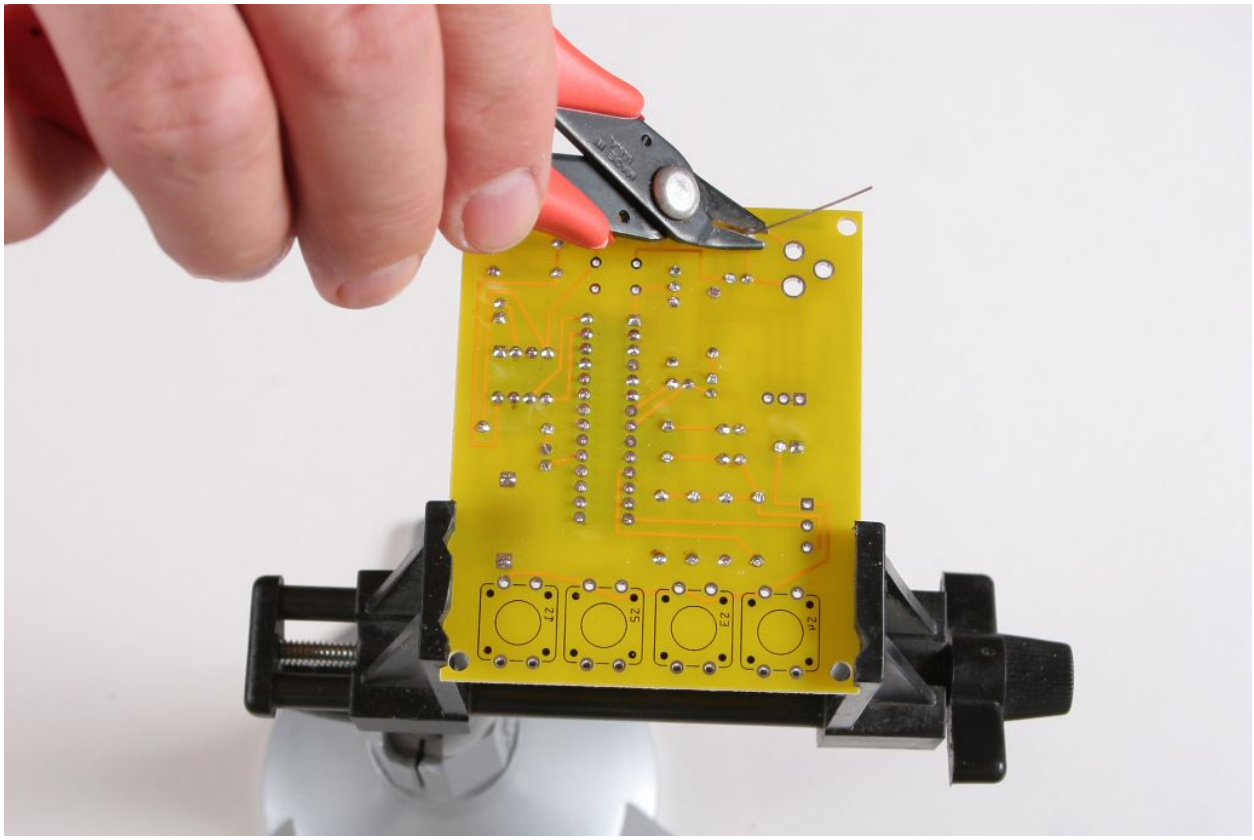


Place the 100uF electrolytic capacitors on C2. Electrolytic capacitors are polarized, make sure to insert the long wire into the round hole marked with a (+) sign.

Solder



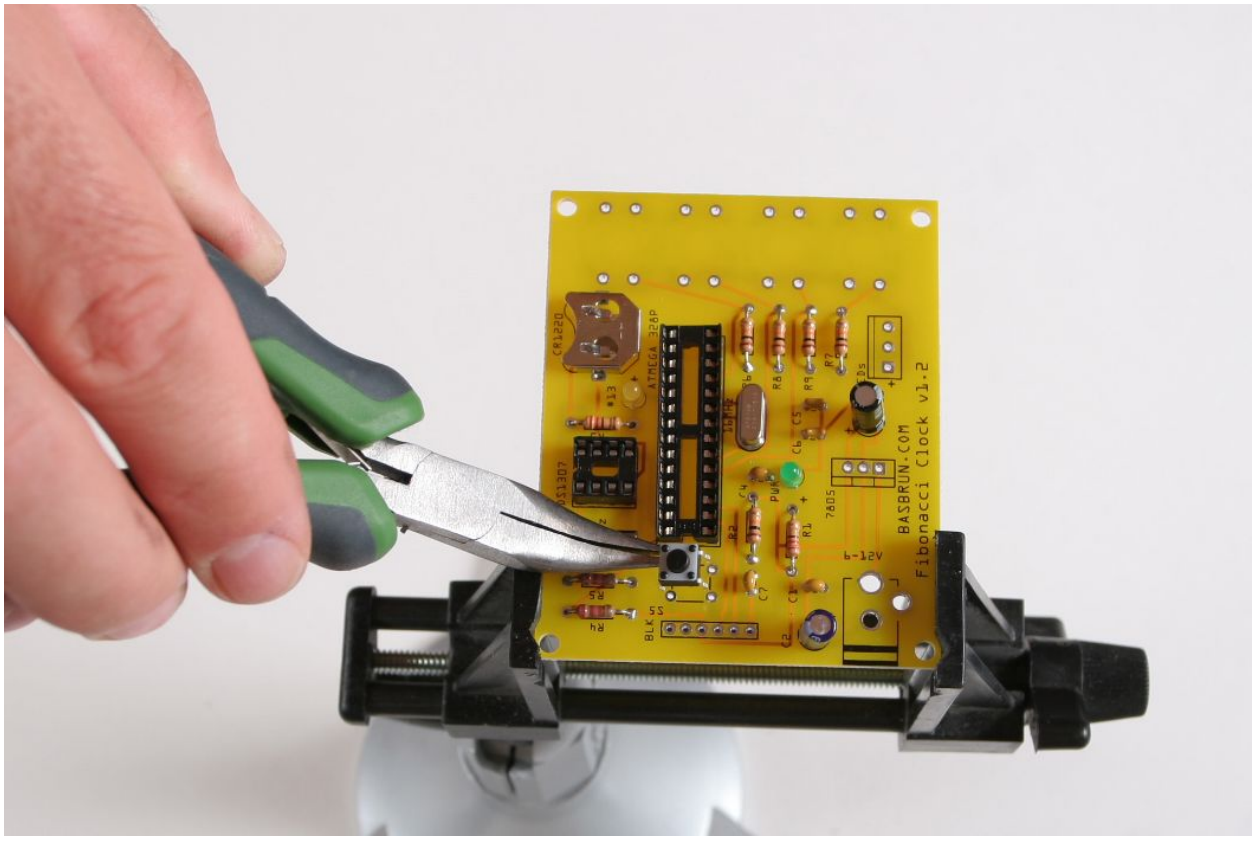
Cut the excess wires.



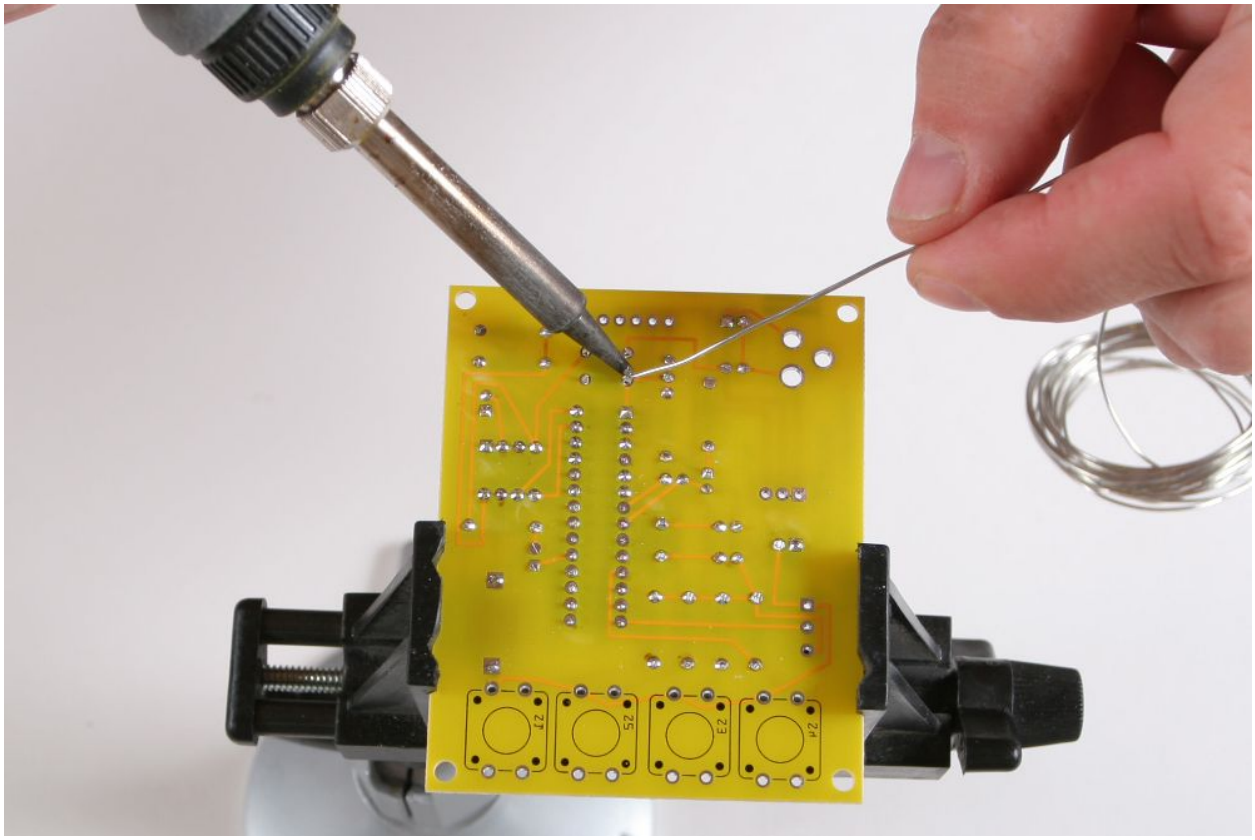
Reset Switch

Parts	Quantity	Image
SWITCH TACTILE SPST-NO 0.05A 24V	1	

Place the reset switch in its location between the Atmega328P DIP socket and the FTDI port. The pins of the switch must be facing the top and bottom sides of the board.



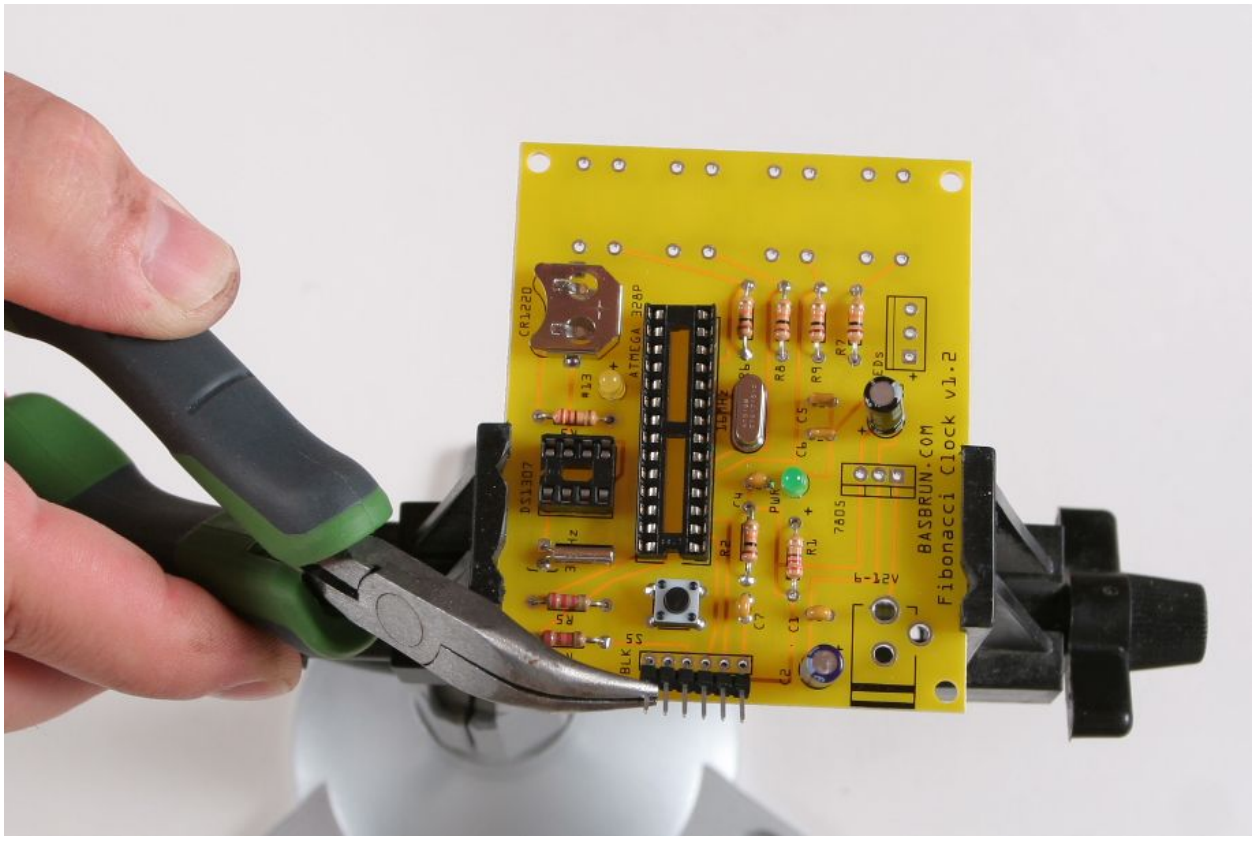
Solder



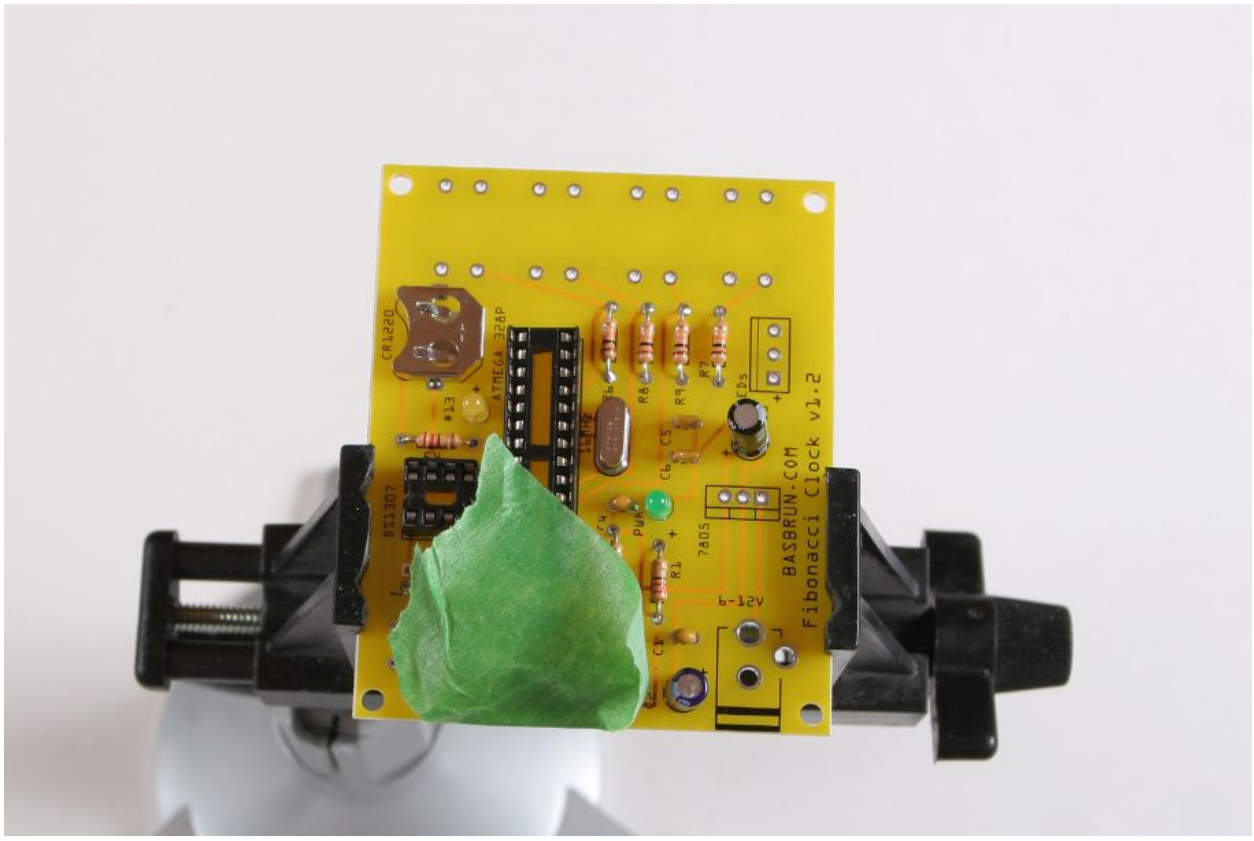
FTDI Port

Parts	Quantity	Image
CONN HEADER 6POS .100 STR TIN	1	

Place the male header with the long pins facing up.



Secure with tape.

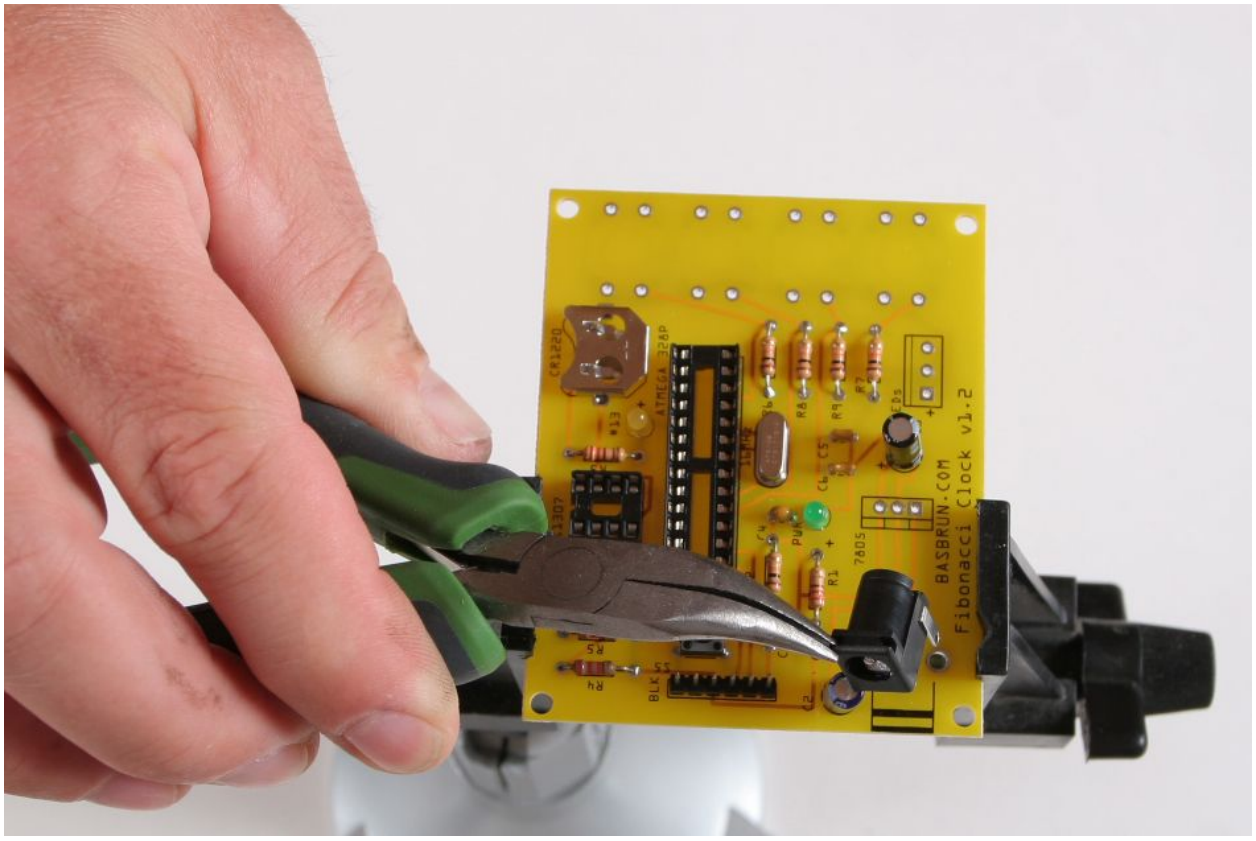


Solder the male header.
[NO IMAGE]

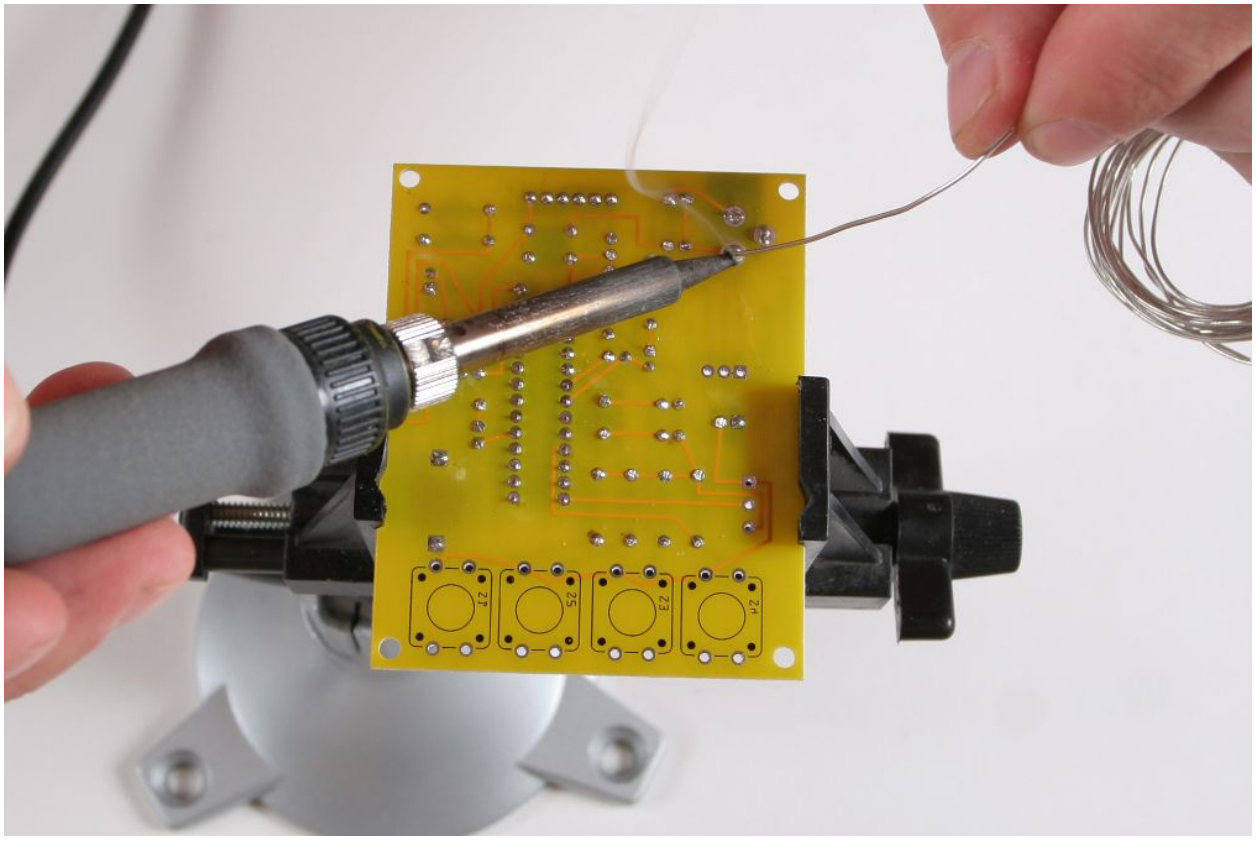
Power Connector

Parts	Quantity	Image
CONN POWER JACK 2.1MM	1	

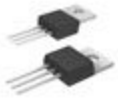
Place the power connector in its location and secure with tape if necessary.



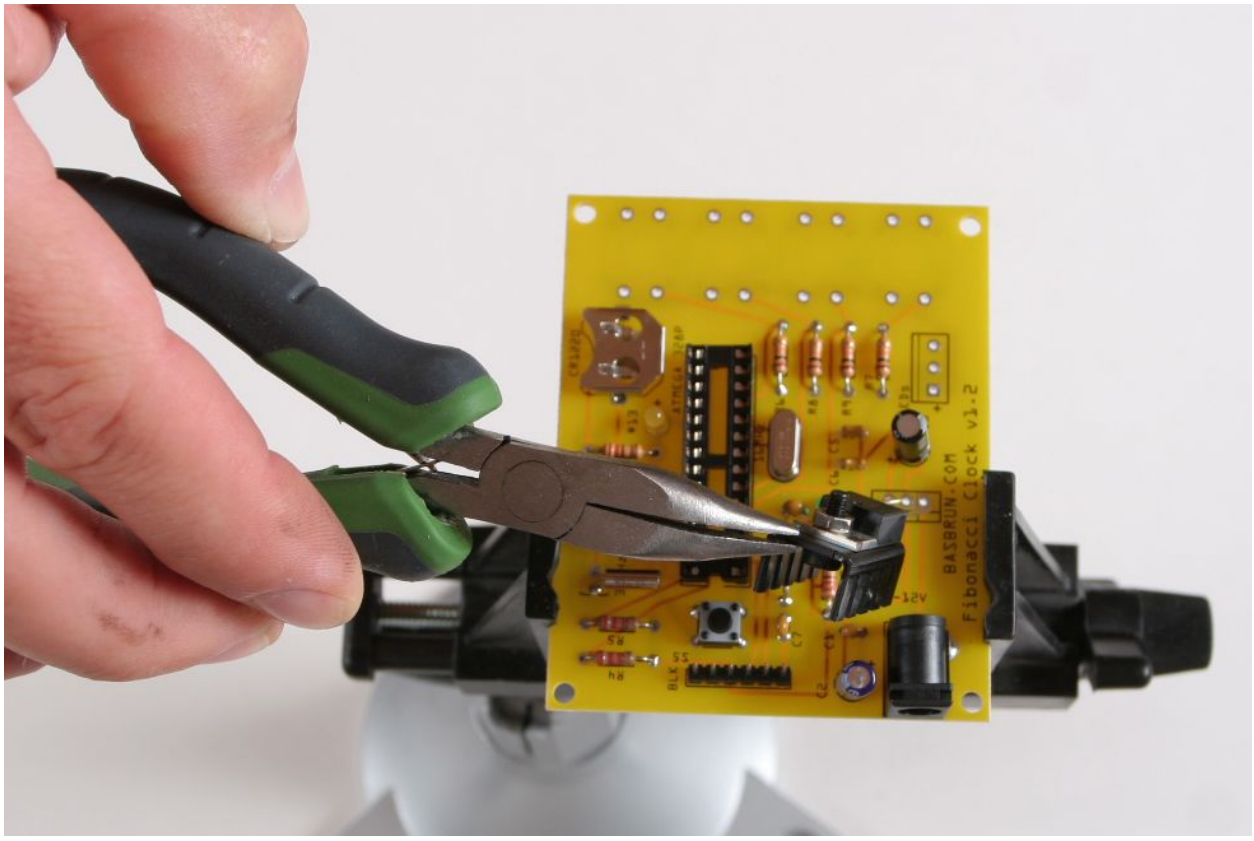
Solder



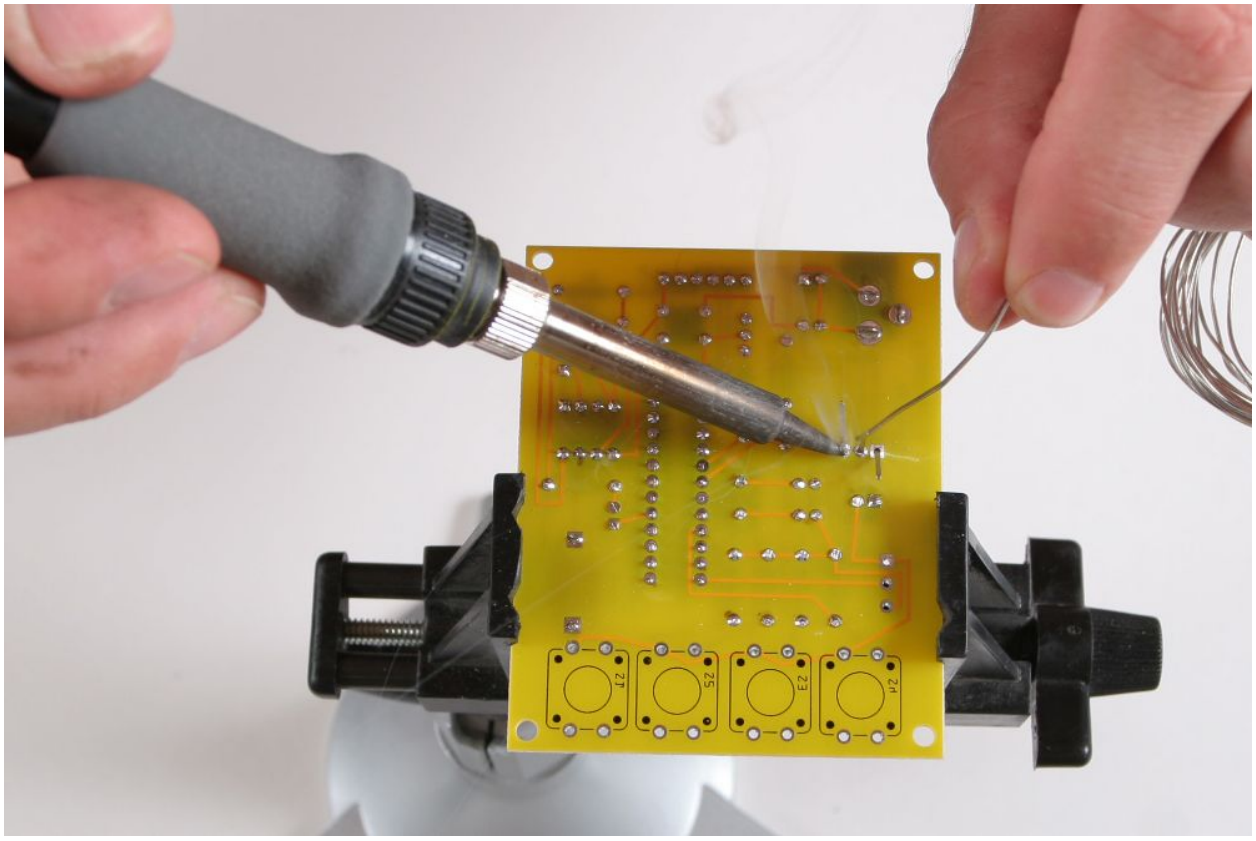
Voltage Regulator

Parts	Quantity	Image
HEAT SINK TO-220 .375" COMPACT	1	
IC REG LDO 5V 1.5A TO220	1	

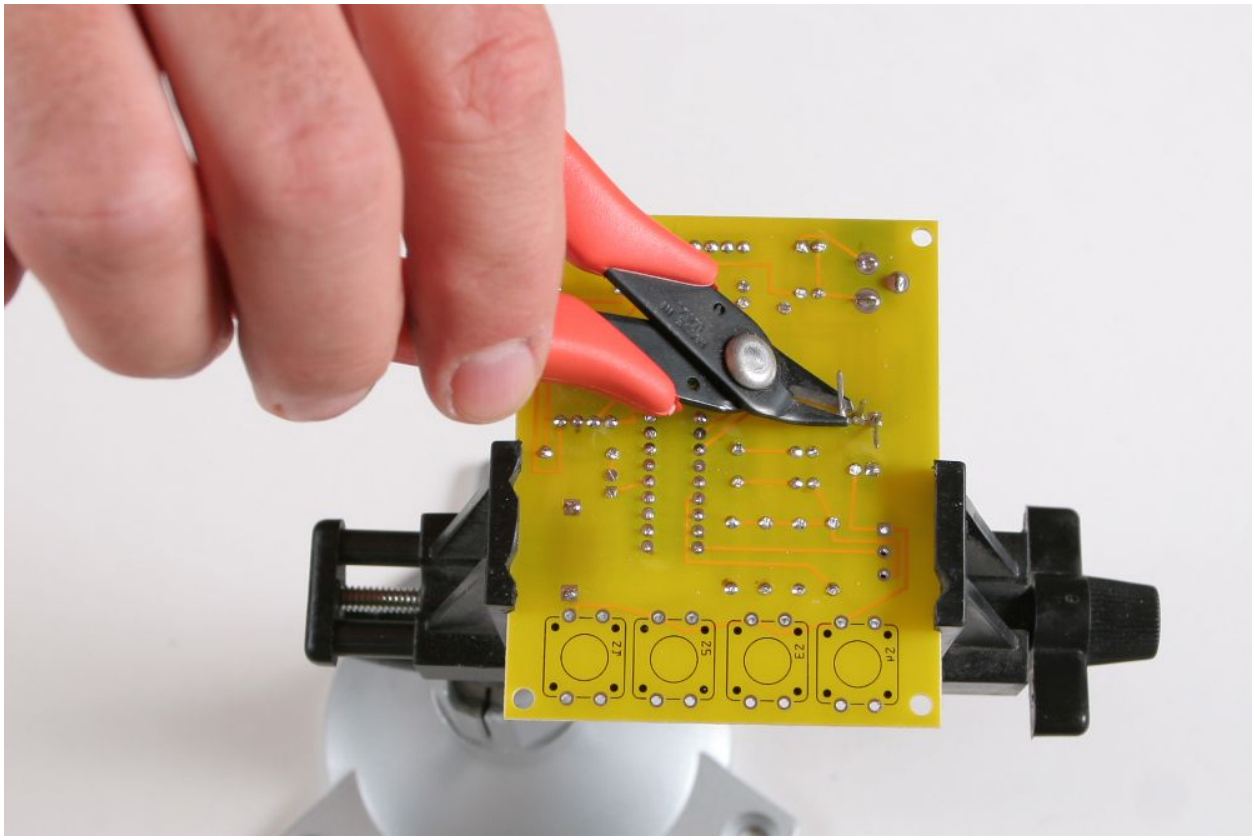
Place the voltage regulator and heat-sink assembly in the place marked 7805. The heat-sink should face the power connector.



Solder



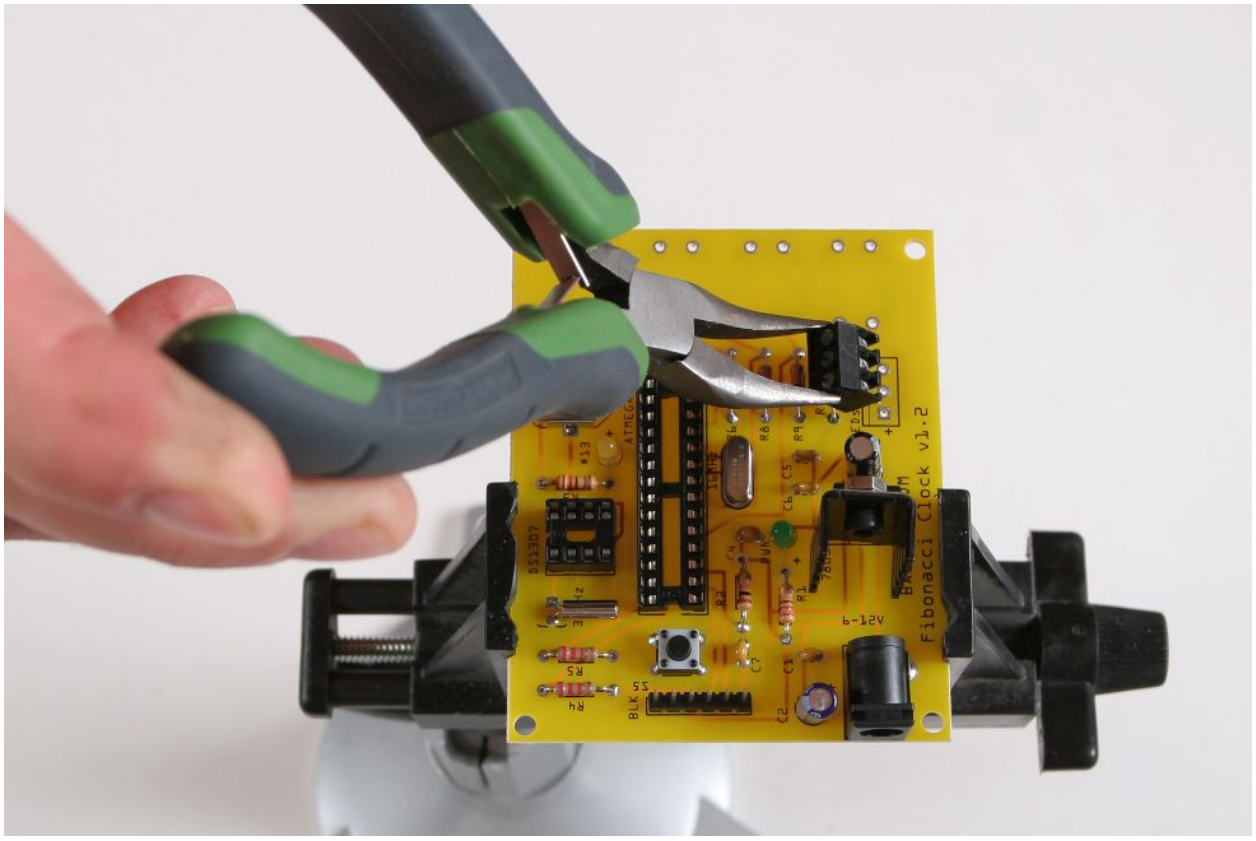
Cut the excess wires.



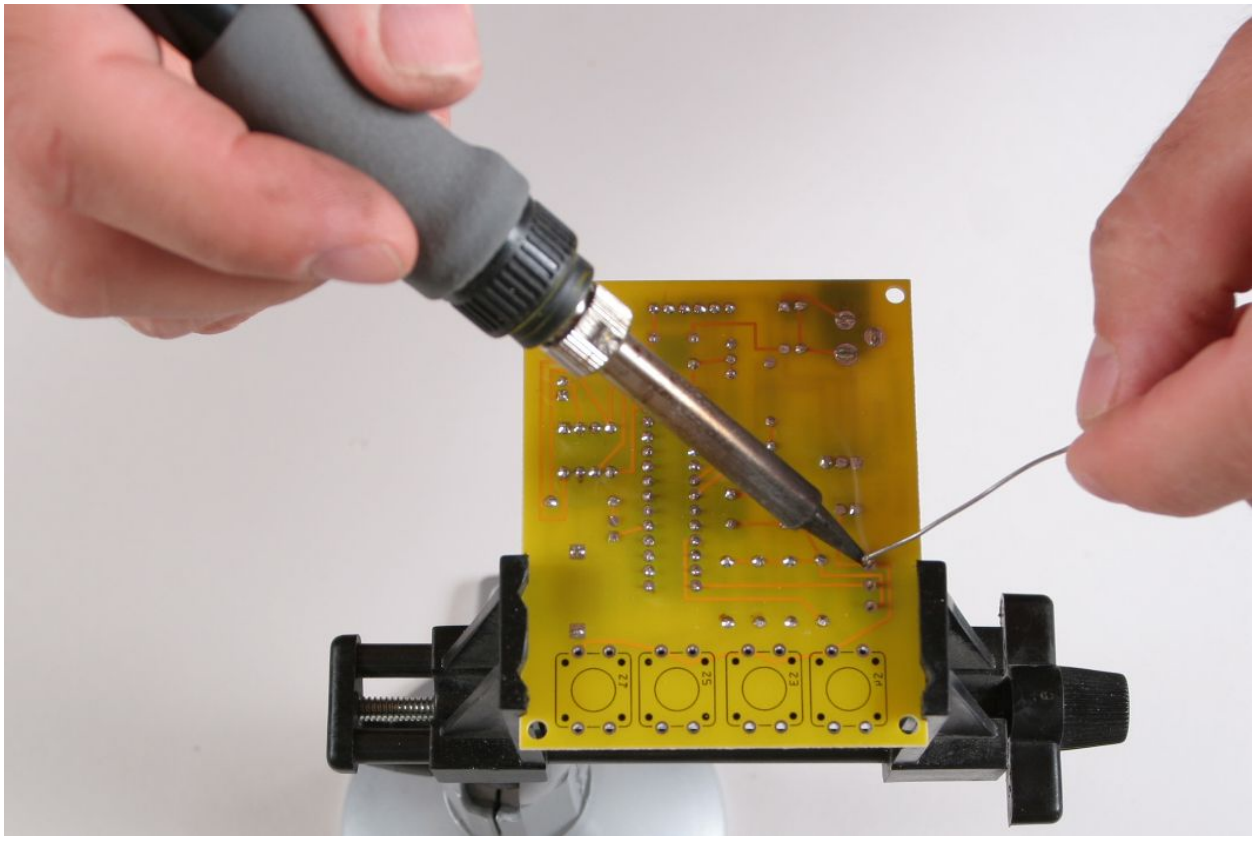
Terminal Block

Parts	Quantity	Image
TERMINAL BLOCK 3.5MM 3POS PCB	1	

Place the terminal block on the board with the beveled side facing the bottom edge of the board. Secure the part with tape if necessary.



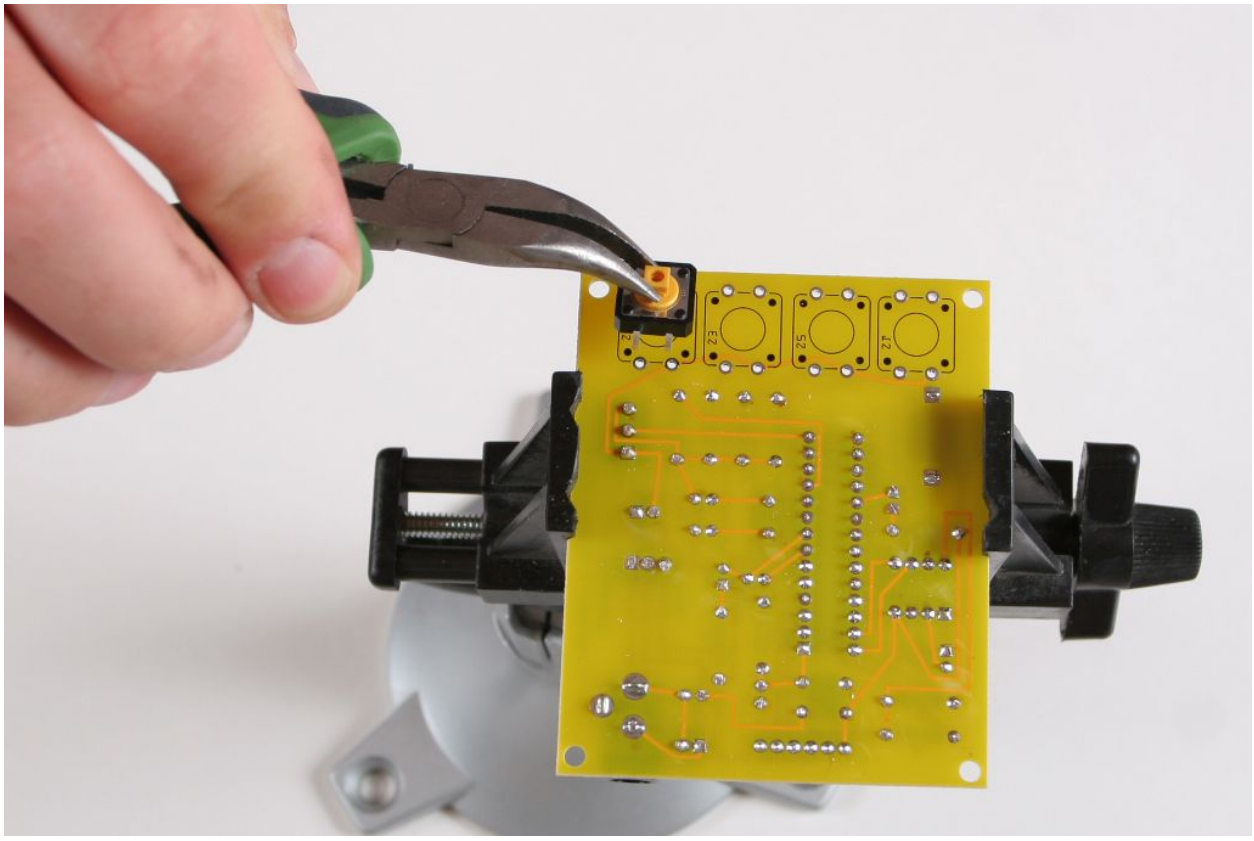
Solder

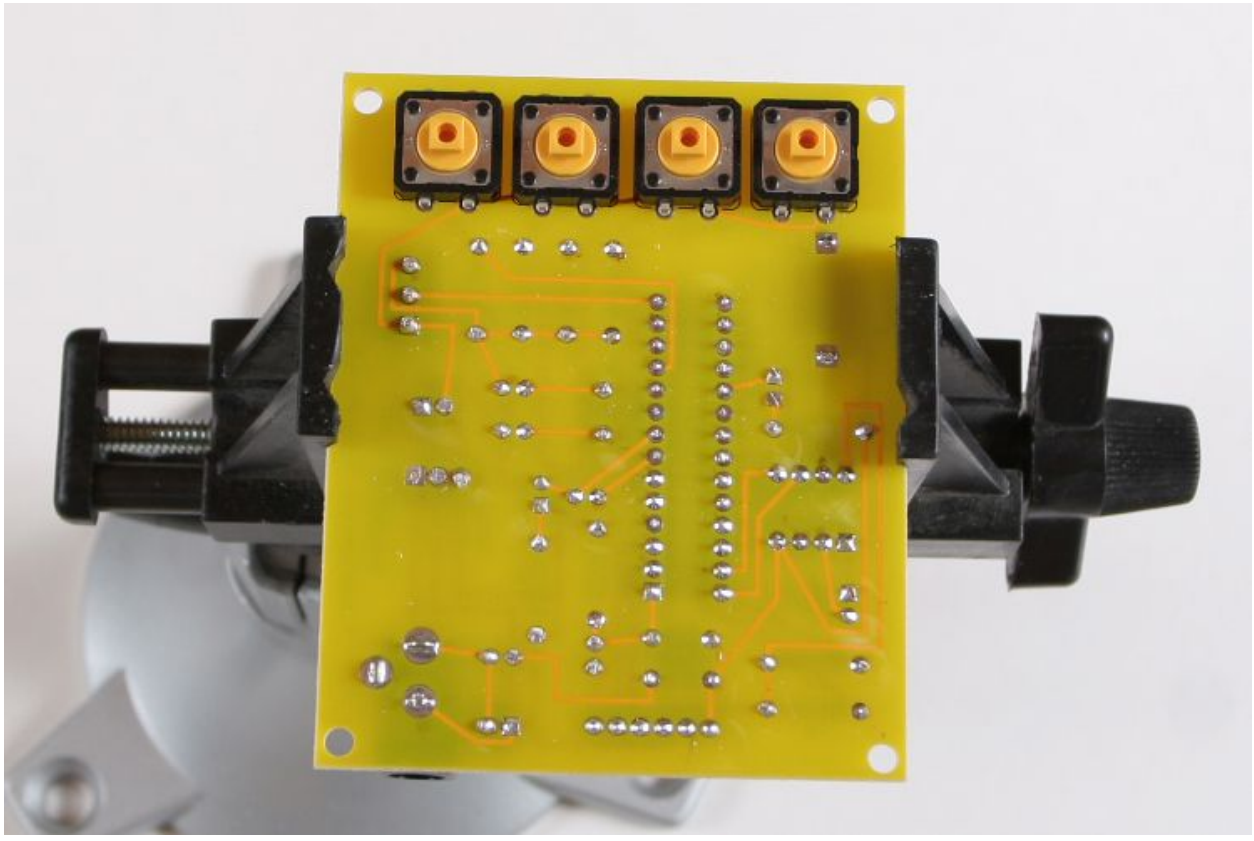


Clock Buttons

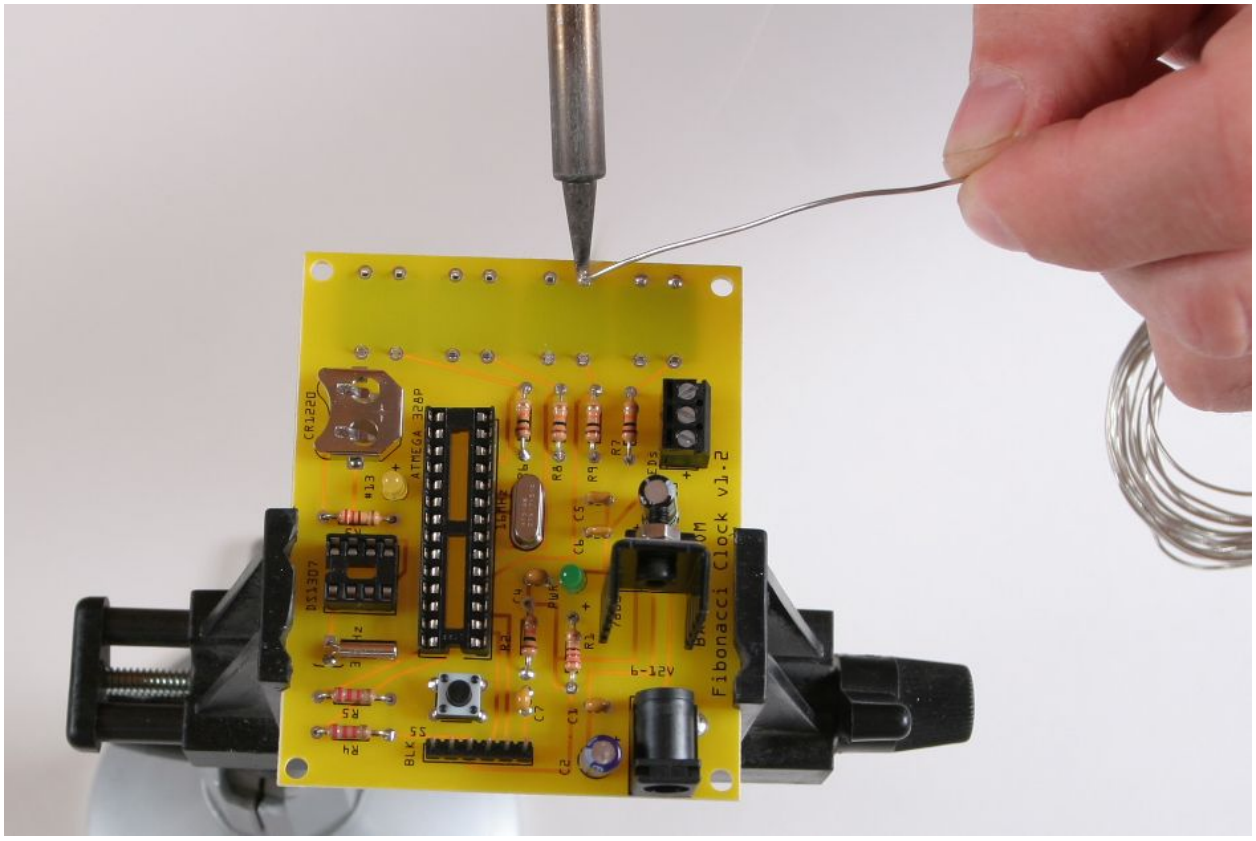
Parts	Quantity	Image
SWITCH TACTILE SPST-NO 0.05A 24V ***May be all black in your kit***	4	

Turn the board on the other side and place the four buttons in their location. They should snap in place.





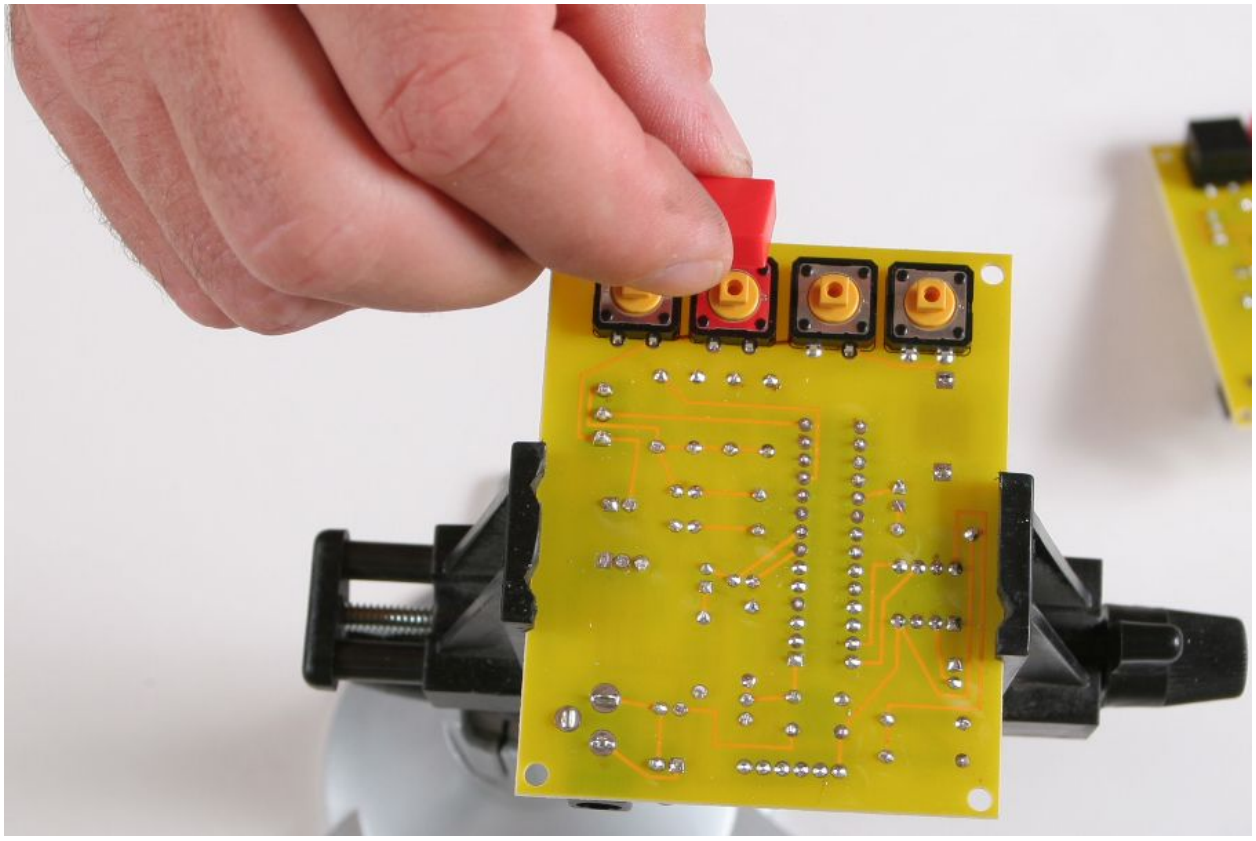
Solder

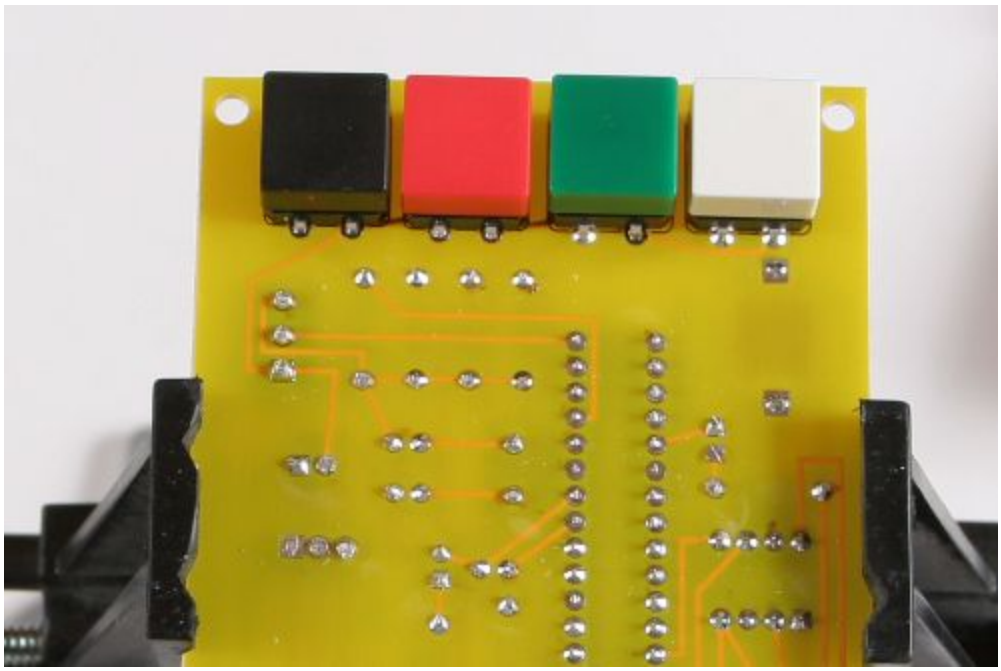


Buttons caps

Parts	Quantity	Image
CAP TACTILE SQUARE RED	1	
CAP TACTILE SQUARE WHITE	1	
CAP TACTILE SQUARE GREEN	1	
CAP TACTILE SQUARE BLACK	1	

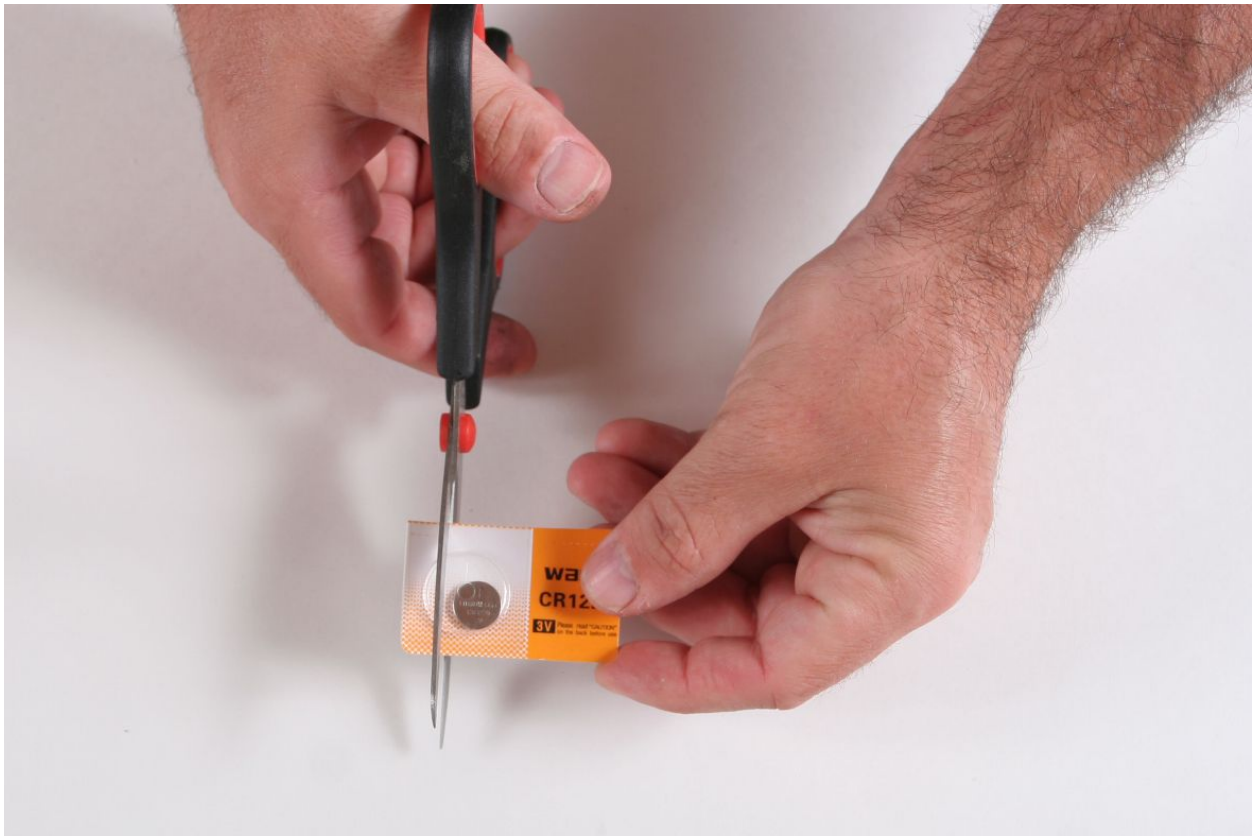
Place the button caps as displayed on the following picture. *Some kits come with button caps with a pin in the center of the socket. You may have to cut it if the button in your kit don't have a center hole.*

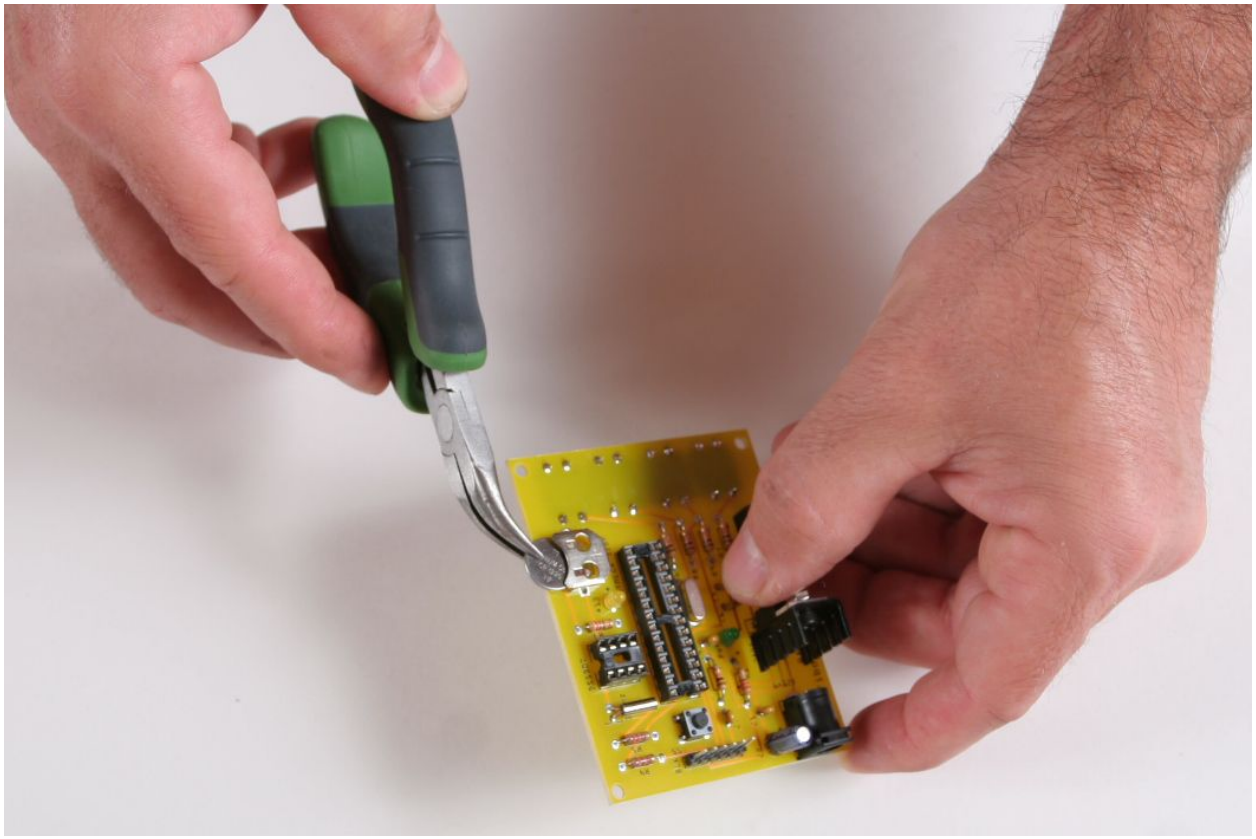


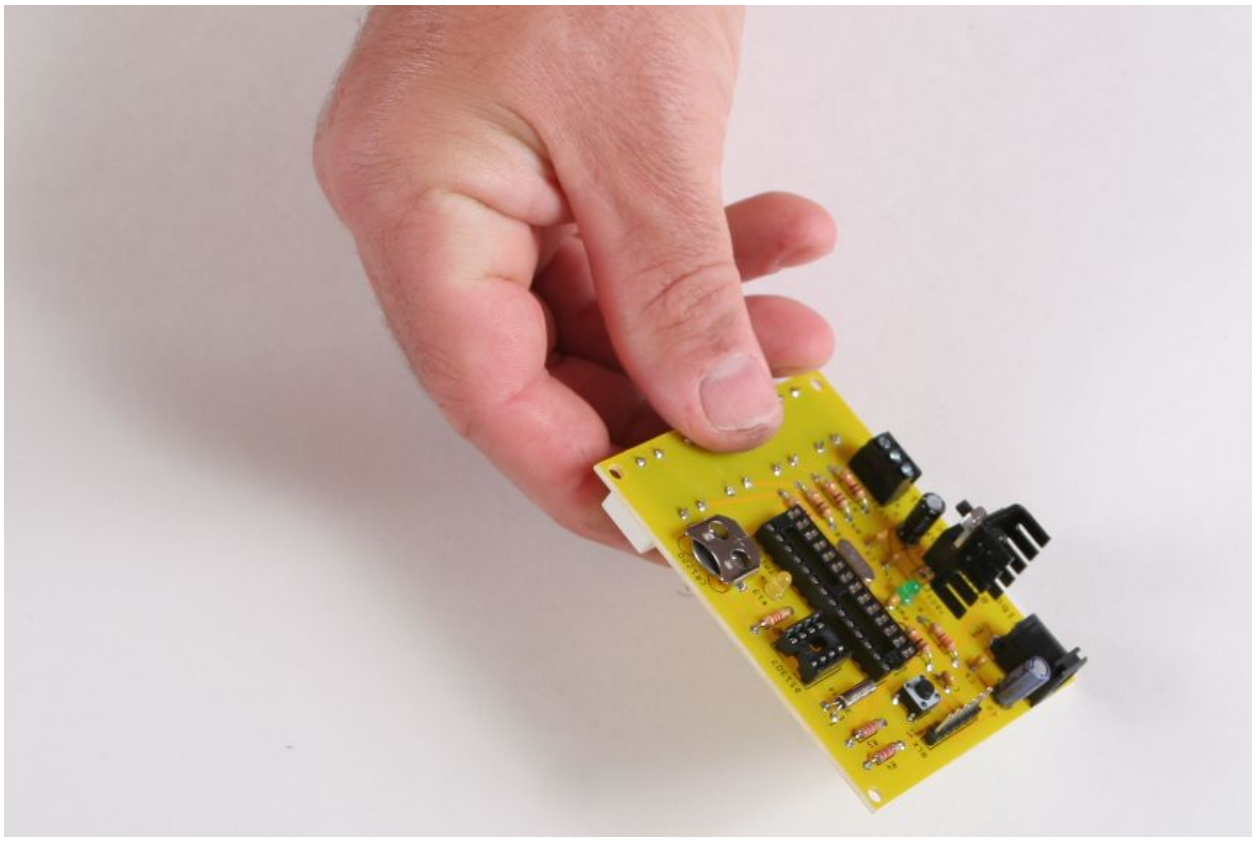


Coin Battery

Insert the coin battery in the battery holder.

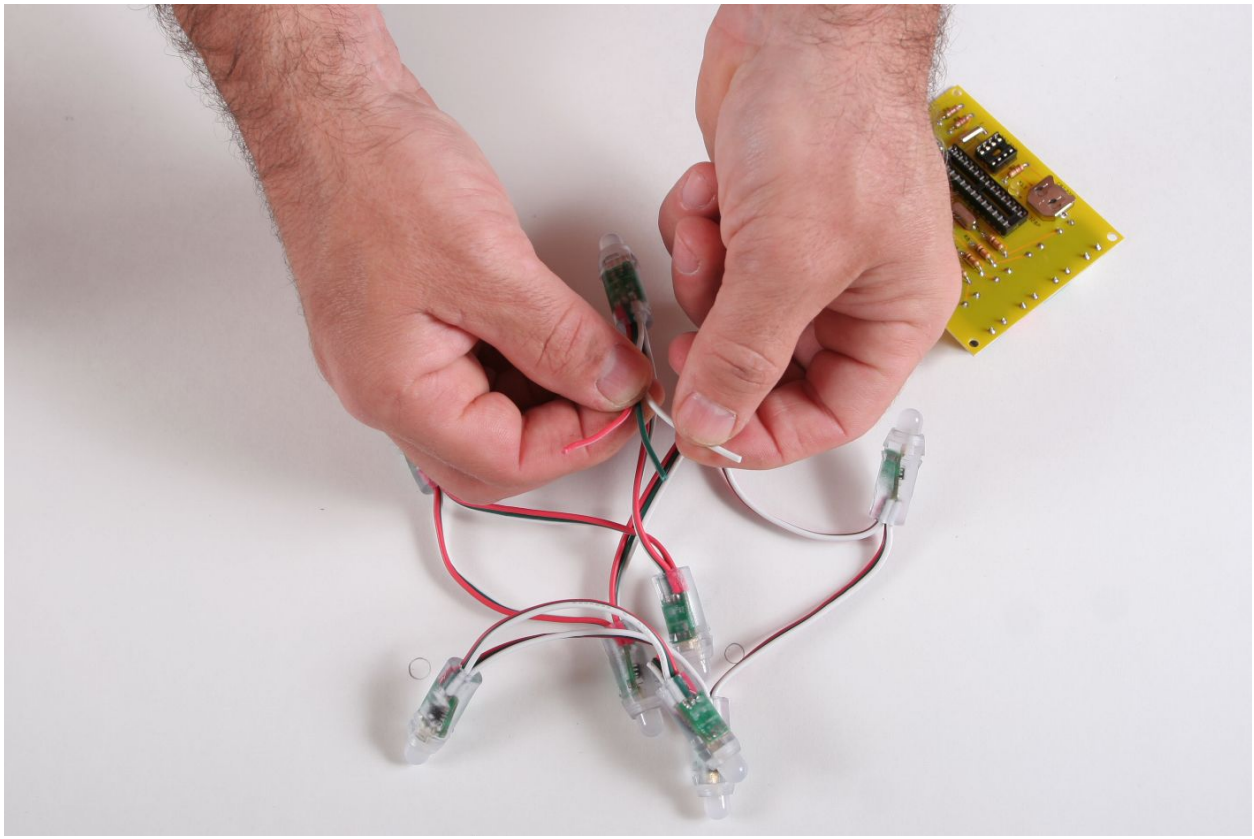






LED Pixels

Separate the three wires of the LED strip.



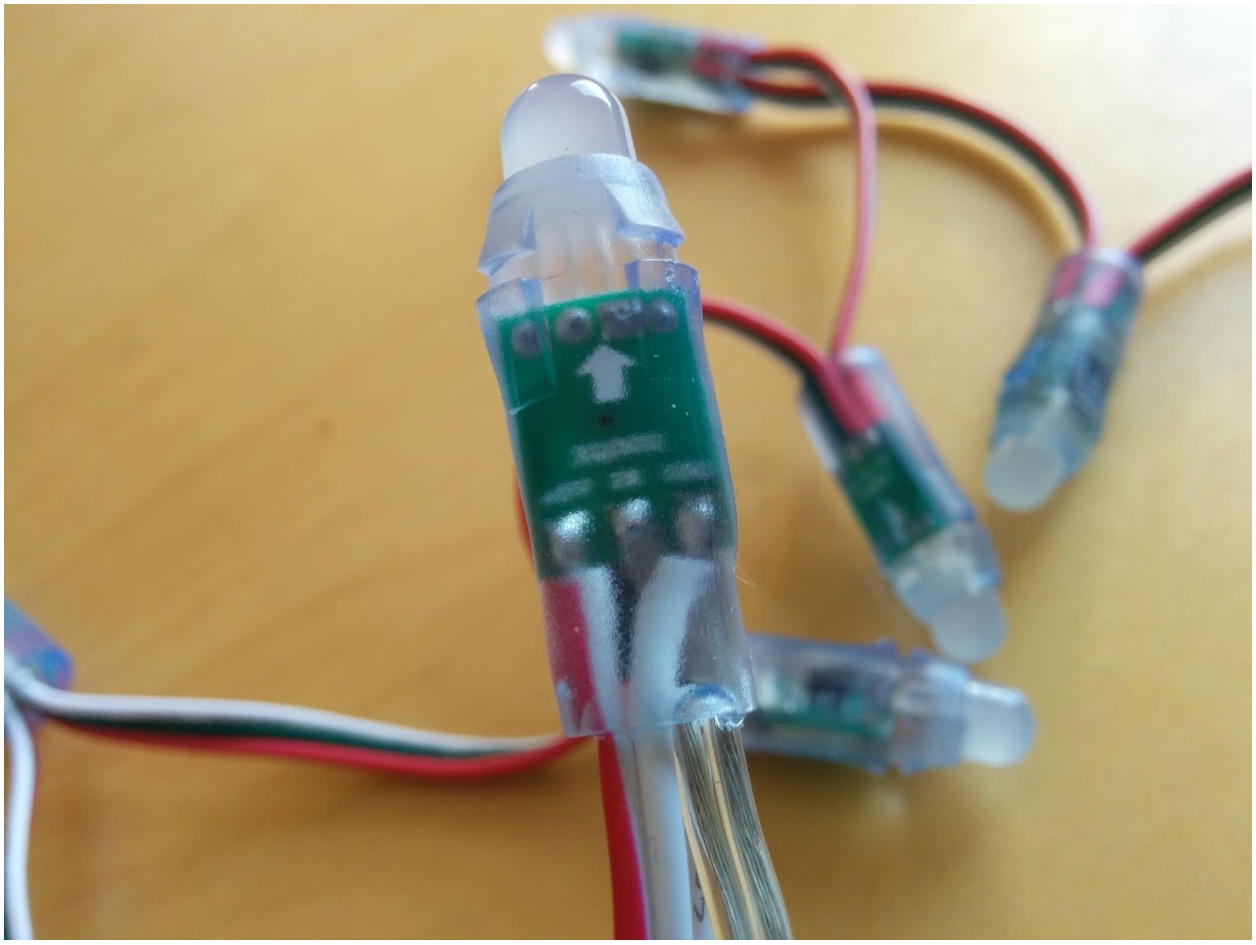
Strip the wires on approximately 5mm.

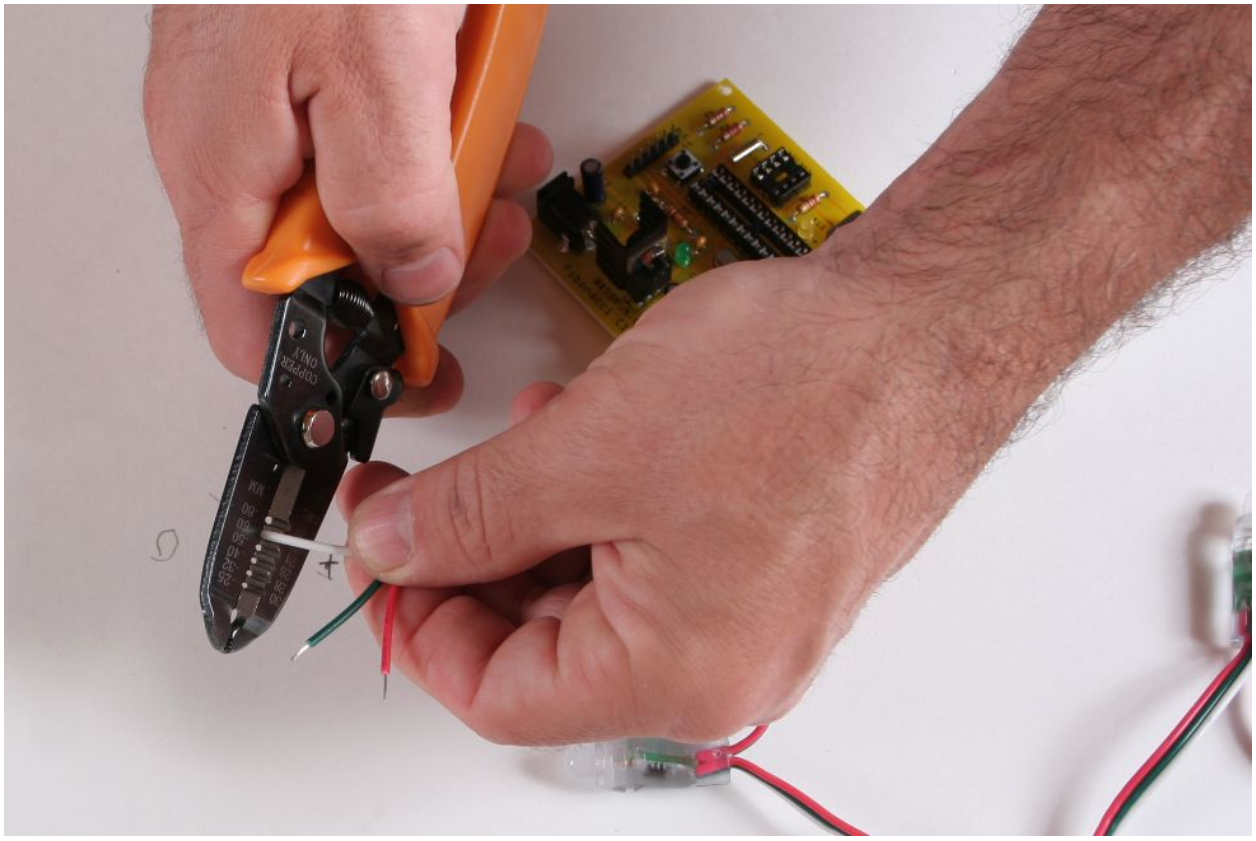
Important note **Some kits will ship with a slightly different LED strip. The green wire being replace by two transparent wires.**

One of the transparent wires replaces the green wire while the other is connected with the white wire on the ground.

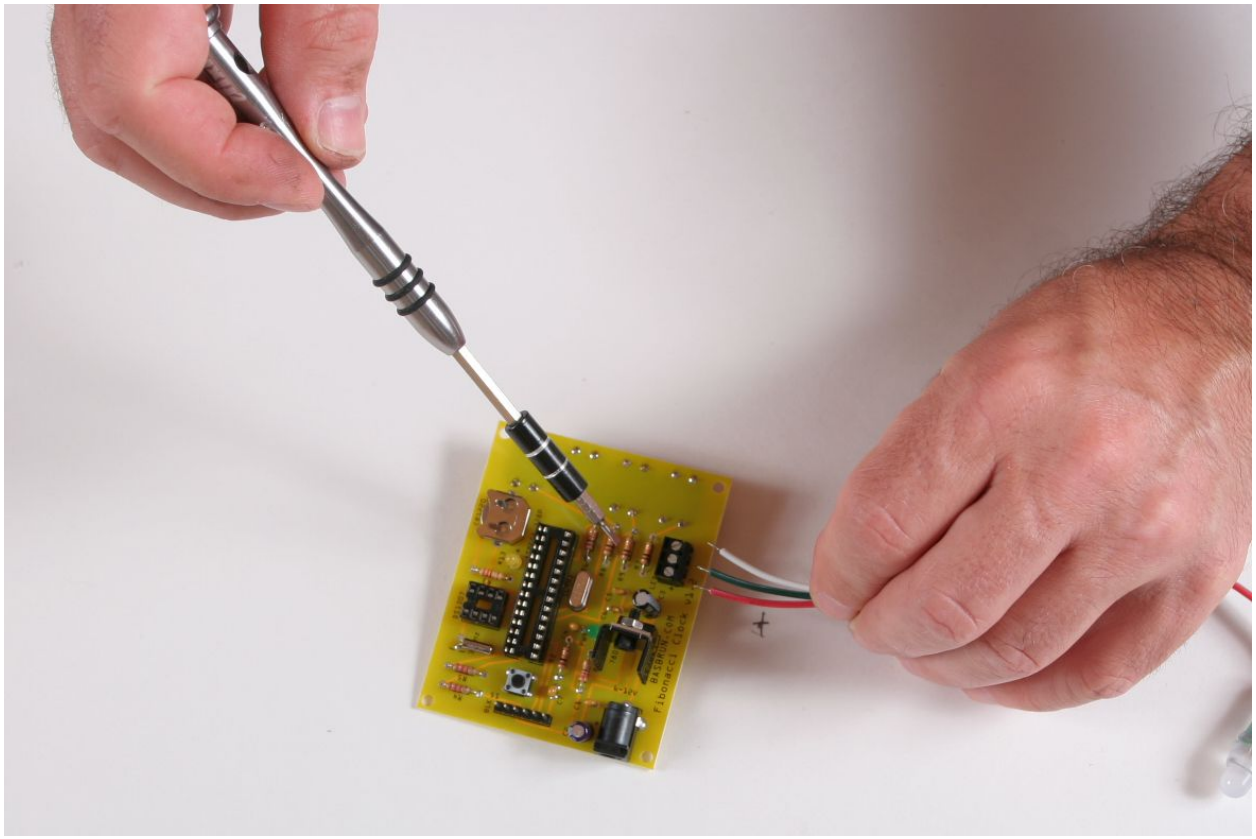
You should use the transparent wire soldered to the center pin of the LED circuit board. If you are not sure what wire this is use a voltmeter to test connectivity between the transparent wires and the white wire. You should connect the one not connected to the ground (white).

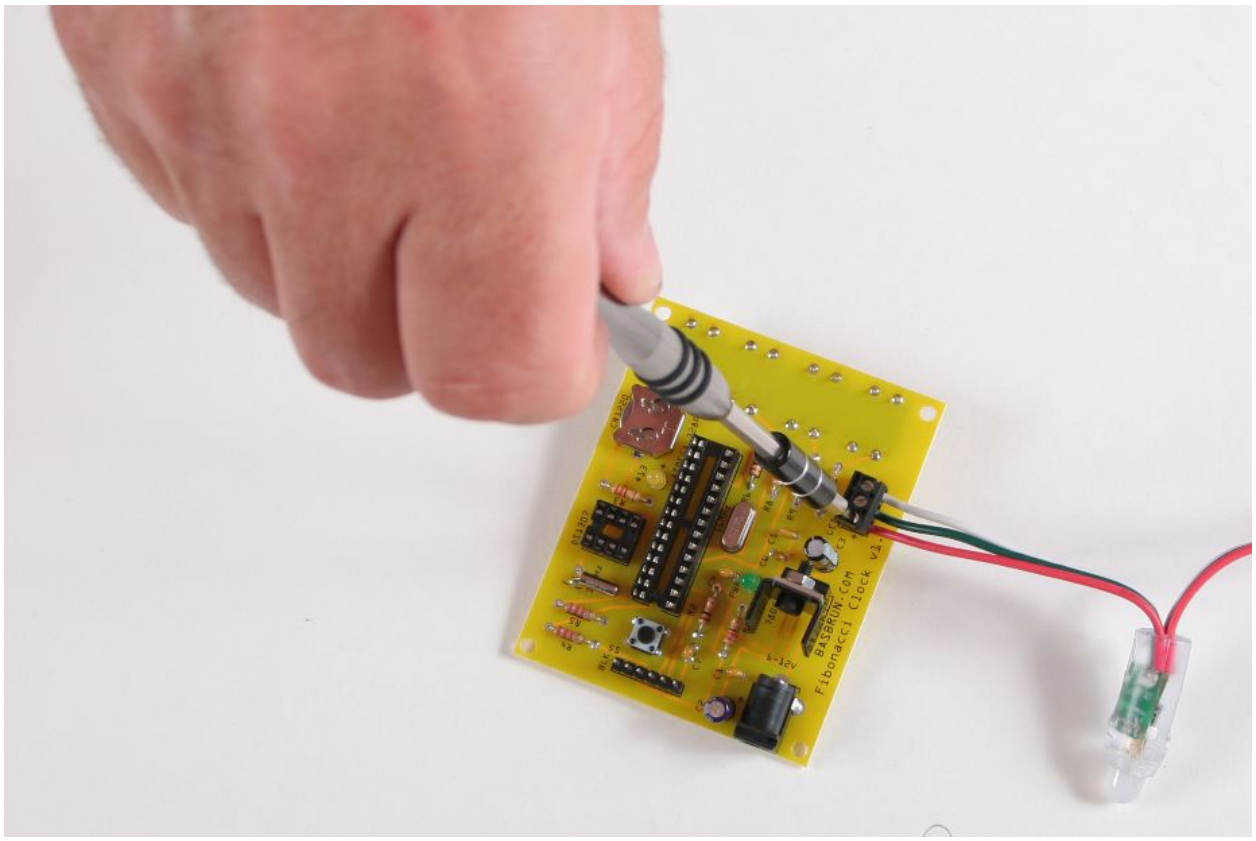
An LED strip with two transparent wires. Uses the one connected to the center pin of the circuit board (IN). The other is connected to the ground with the white wire and can be ignored.





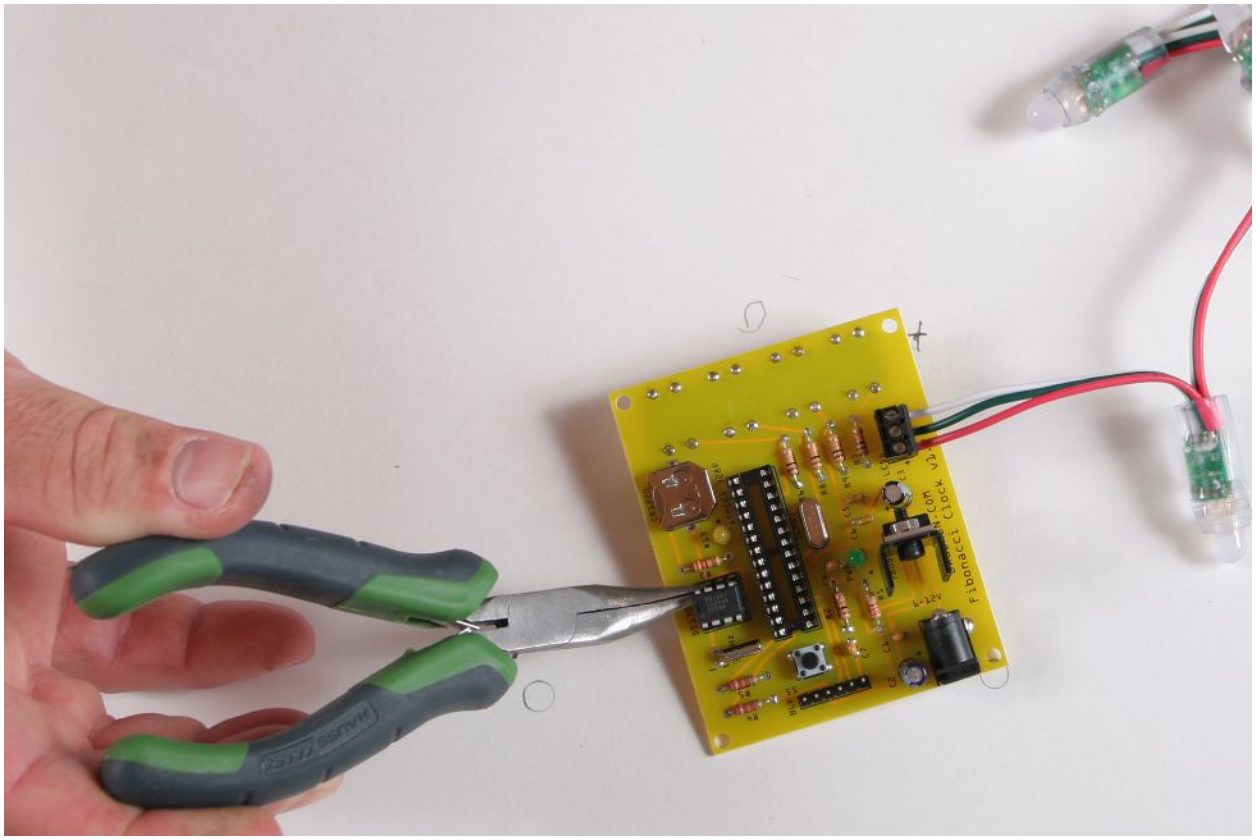
Screw the wires in the block terminal. Insert the red wire in the terminal marked with a (+) sign.

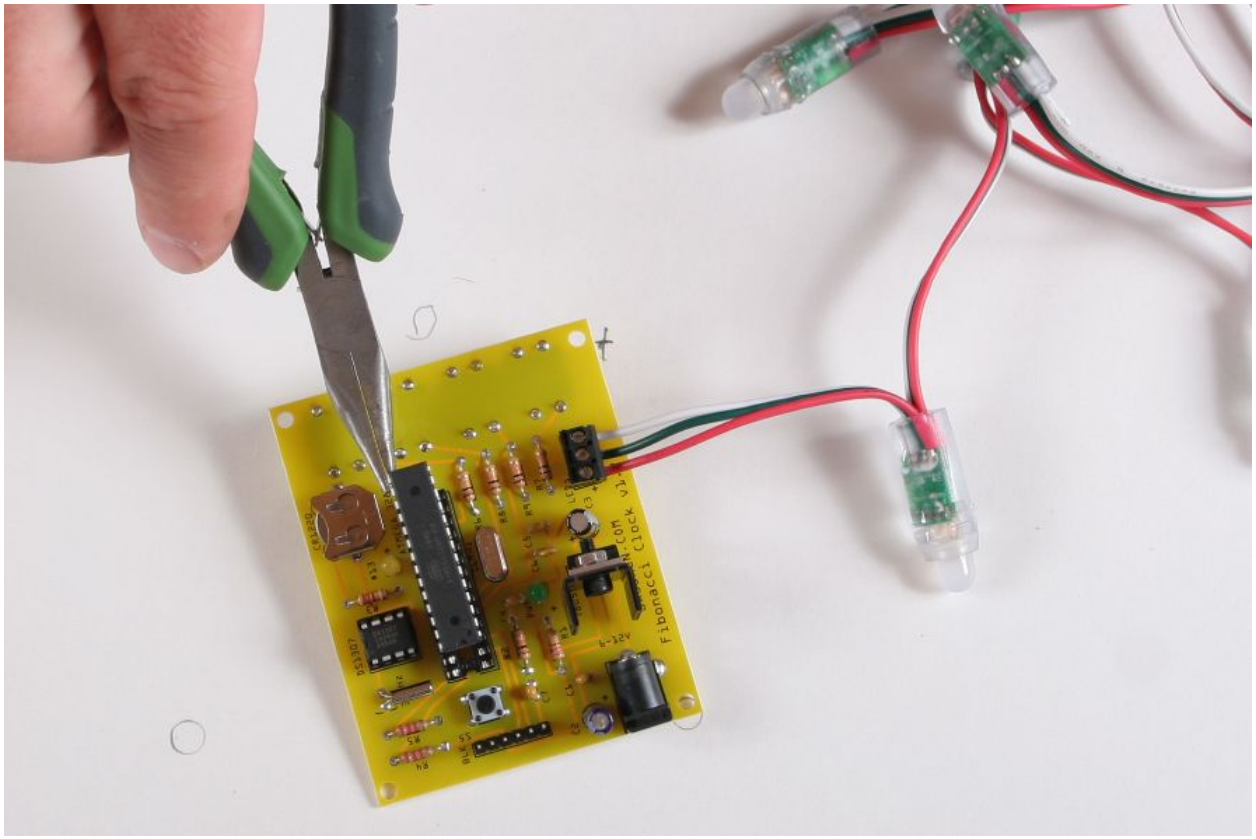

BASRUN.COM
Fibonacci Clock v1

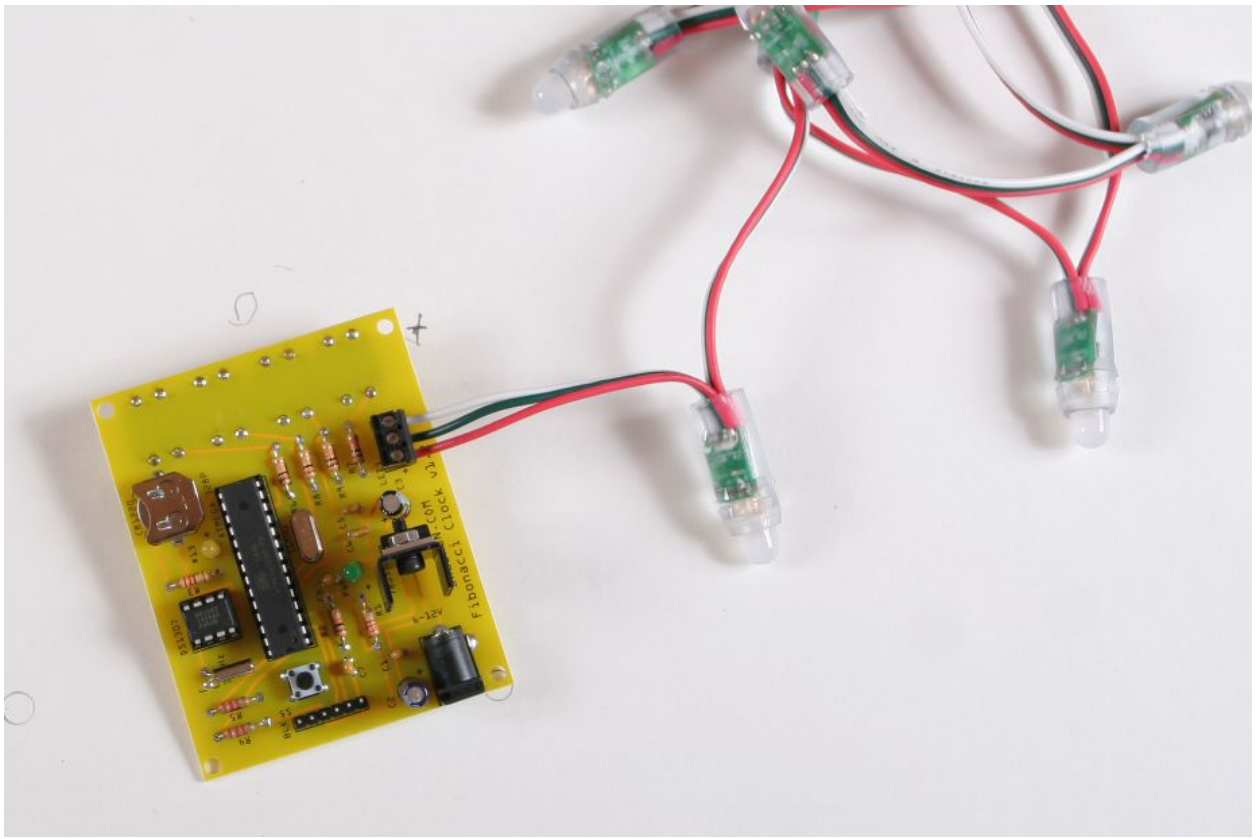

BASRUN.COM
Fibonacci Clock v1

ICs

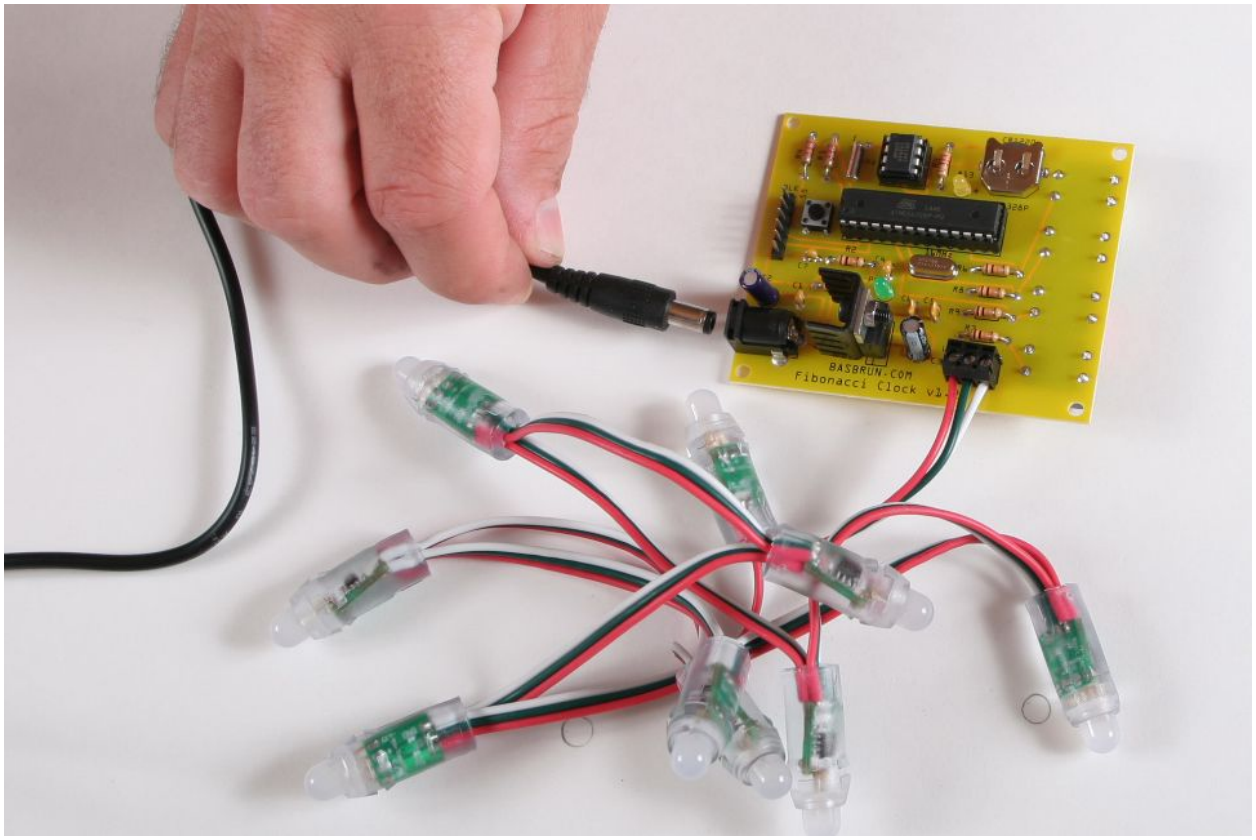
Insert the DS1307 real time clock IC and the Atmega328P microcontroller in the two DIP sockets. Make sure the dents at the end of the ICs are aligned with their equivalent on the DIP socket.

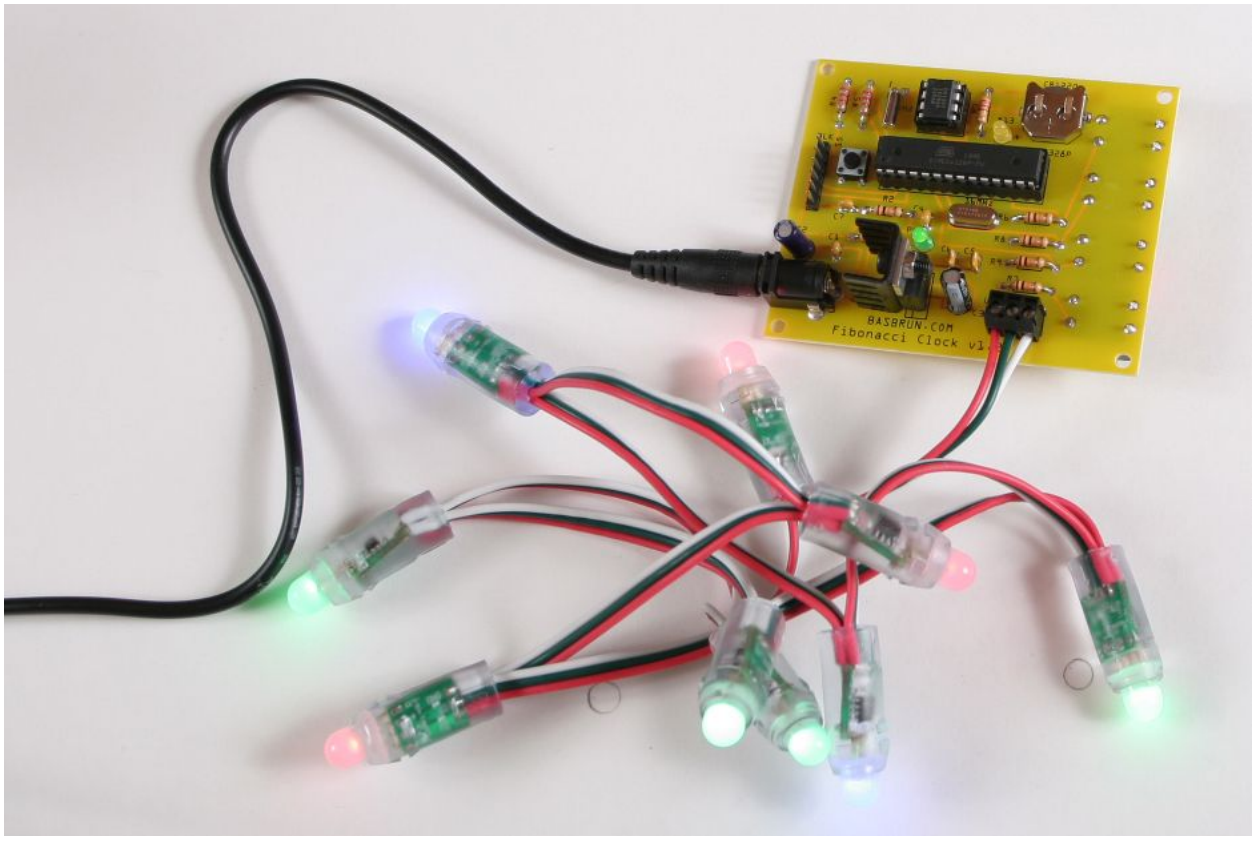



Fibonacci Clock v1


Fibonacci Clock v1

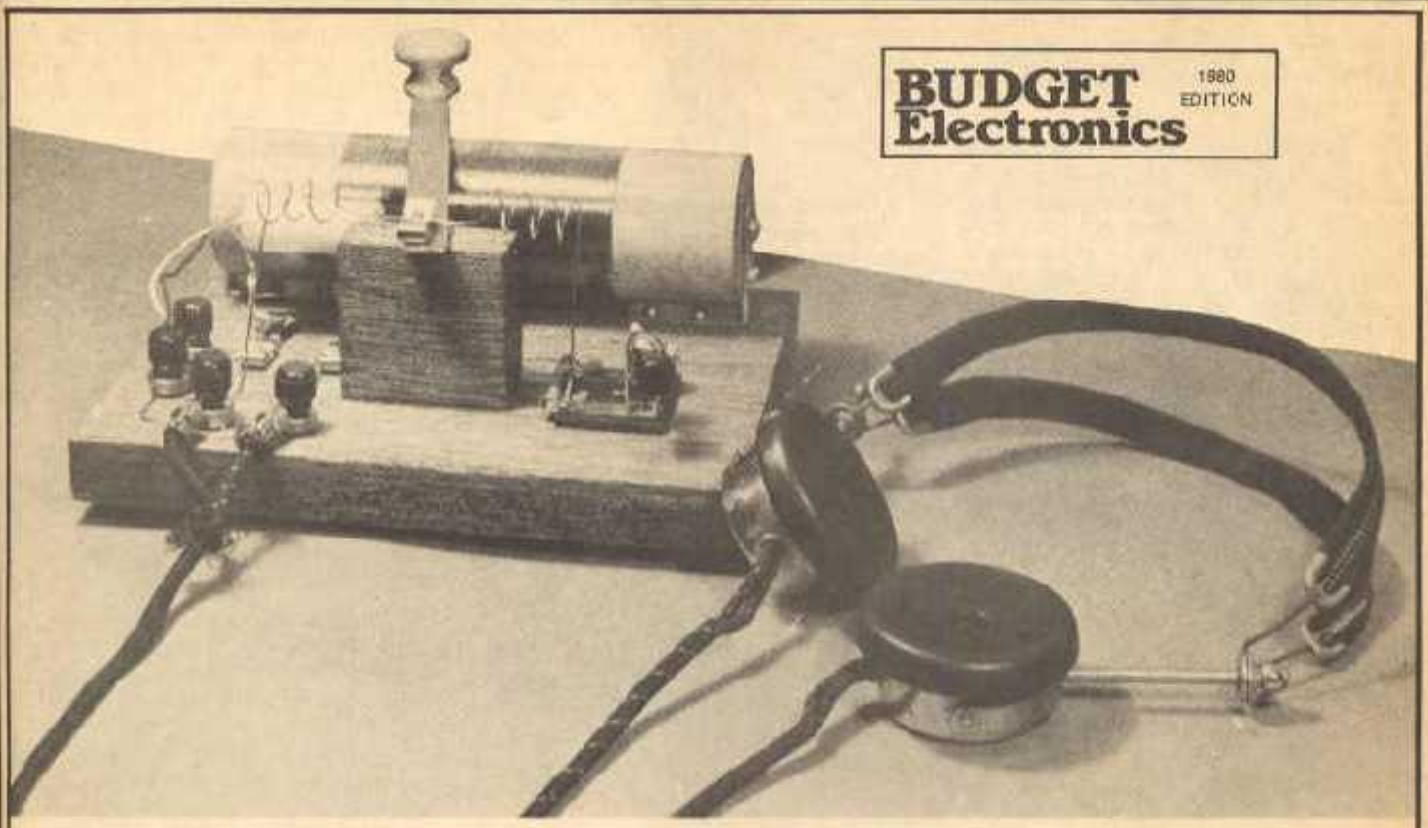
Connect the 6V power to test the circuit.





You're done.
Bravo!

#



CLASSIC CRYSTAL

A low-cost crystal set that will enrich any collection.

by George Campbell

YOU CAN RELIVE THOSE EARLY days of broadcasting by building this authentic reproduction of a crystal radio from the early part of this century. This afternoon's project uses the same components, mostly hand-built, that your grandfather used to use, and costs less than $10 to build.

The heart of this receiver is an authentic crystal detector, using a natural galena crystal and a cat-whisker. It is available by mail order (see parts list for ordering information). The coil and the condenser are also handmade, using original techniques from the early '20s.

Old-Style Components. Begin by laying out the wood base according to the drawing. You may use any wood you choose, as long as it's dry and well-seasoned. Drill 3/32-inch pilot holes for the screws and 5/32-inch holes for the binding posts. Counterdrill the binding post holes ¾-inches deep with a ⅜-inch bit. Glue the slider support block to the base, then give the entire base two coats of shellac, the original finish for most of those old-time sets.

While the base dries, make the coil and condenser. The coil is wound on 4 inches of a 7-inch section of a rolling pin. Remove the handle from one end and withdraw the iron rod that passes through the rolling pin. Cut the rolling pin to length, and put the nylon bearing from the unused end of the pin in the hole in the 7-inch section.

Drill 1/16-inch holes completely through the wood, 1½ inches from each end. Use the original rod from the pin as an axle, and carefully wind a single, smooth layer of No. 25, coated copper magnet wire between the holes. The coil must be tight, with no overlaps. Pass the wire ends through the pre-drilled holes to anchor the coil, then give it two coats of shellac.

The crystal/whisker act as a rectifier diode to detect audio from the tuned LC circuit.

Next, cut the parts of the condenser as shown in the diagram. On a hard surface, stack the strips together, beginning with the Kraft paper, then waxed paper, then foil, waxed paper, foil, waxed paper, and Kraft paper.

The roll-your-own capacitor is made of metal foil and paper—just like the old days.

CLASSIC CRYSTAL

The L-shaped ends of the foil should be at opposite ends of the stack, and protrude from opposite sides. Make sure the foil is centered between the layers, then bind the edges of the condenser together up to the foil with masking tape. Roll the bundle up tightly. Cover the completed condenser with masking tape and crush the protruding foil into wire-like leads.

Make the slider from a ⅛- by ½-inch brass strip, 6 inches long. Round the corners with a file and drill a 5/32-inch pivot hole, 1 inch from one end and a 3/16-inch hole for the knob at the other. Solder a 3-inch piece of No. 12 bare copper wire to the underside of the slider to make contact with the coil.

Assembly. Mount all components on the base, using No. 5 x ½ inch brass, round-head wood screws. Place a fiber washer under the tuning slide pivot to reduce friction. Locate the tuning coil so that the holes in the nylon bushings line up with the top holes in the brass L-brackets. Use a No. 8 wood screw here. Attach the rubber feet to the bottom of the base.

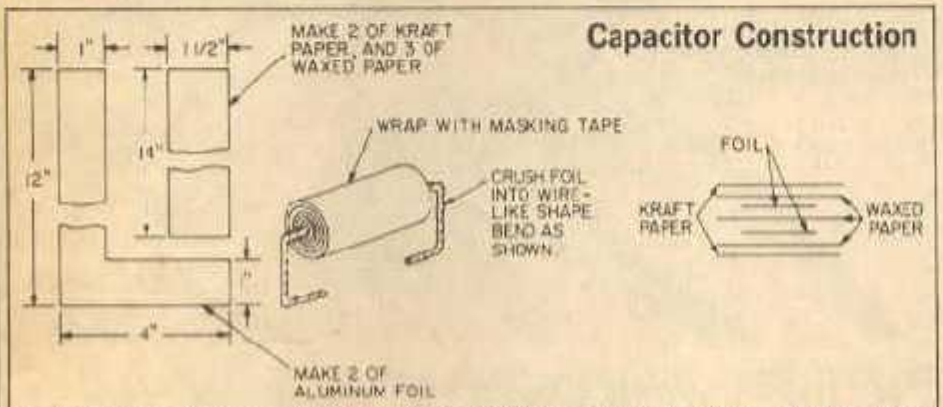
Wire the receiver according to the drawing, using No. 20 copper wire. Remove the insulation and shellac from the top of the tuning coil with fine sandpaper, and bend the tuning slide slightly to provide good contact with the bare wire of the coil.

To use your crystal set, connect the antenna binding post to a good anteanna (one side of a TV lead-in works well). The ground post must be connected to a good ground, such as a cold water pipe. The headphones used with this set must be the high impedance type. Normal stereo headsets will not work with this primitive receiver. If you wish, you can attach the center conductor of a shielded cable to binding post J3, the shield to J4, and plug the other end into an amplifier to provide loudspeaker reception and easier listening.

Locate a sensitive spot on the galena crystal by slowly dragging the cat-whisker spring over its surface with the tuning slide centered on the coil. Once a sensitive spot is found, you should hear a local station clearly. Tuning this receiver takes time and patience, so don't give up.

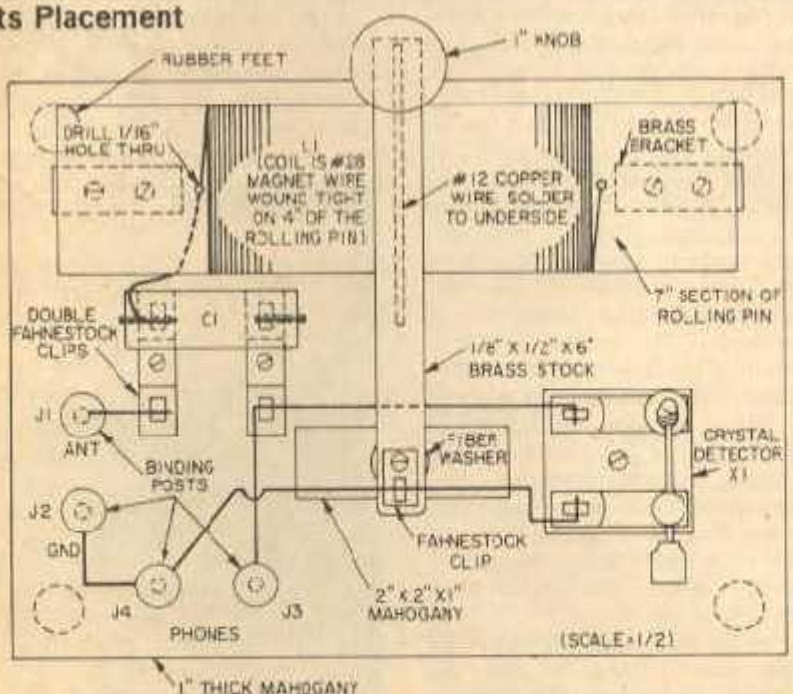
Don't be surprised if you hear more than one station at once. Crystal sets are not very selective. However, careful adjustment of the tuning slide should separate stations reasonably well. And who knows? One day you may even hear a voice from the past. ■

Capacitor Construction

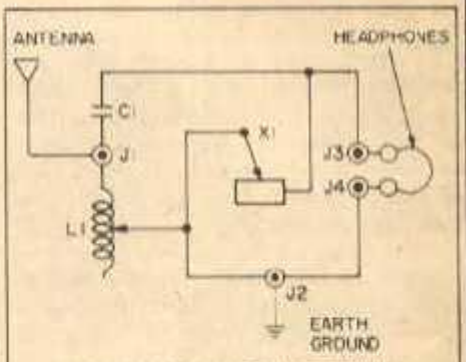


To build the capacitor (C1) cut strips of aluminum foil, kraft paper, and wax paper according to the patterns above. Then interleave the strips as shown above, with the L-shaped ends of foil at opposite ends of the bundle, facing in opposite directions. Bind the edges with masking tape and roll the bundle tightly. Test it with a VOM: if the meter moves and then goes back to infinite resistance then C1 is OK. If not, rebuild C1.

Parts Placement



This parts' location diagram shows the relative positions of all the parts. It is drawn to half-scale, so multiply any distance by two to get the full-scale measurements. The wire connections are the bold lines. Wind the coil for four inches along the rolling pin. The non-functional coils on each side of the wiper in the photograph are for decoration.



PARTS LIST FOR CRYSTAL SET REPLICA

C1—capacitor (see diagram)
J1, J2, J3, J4—birding posts
L1—tuning coil (see text and diagram)
X1—Crystal detector, available from: Johnson Smith Company, 35075 Automation Drive, Mt. Clemens, MI 48043. Order catalog No. 6590. Enclose $1.95 plus $0.60 for postage and handling. Allow 3 weeks for delivery.
Misc.—6- x 8-inch wooden base, 2- x 2- x 1-inch wood slider support block, aluminum foil, wax paper, kraft paper, 150 feet No. 26 coated, copper magnet wire, 3 feet No. 20 copper wire, 3 inches No. 12 copper wire, wooden rolling pin, 2 No. 18 double Fahnestock clips, 1 No. 18 single Fahnestock clip, ⅛- x ½- x 6-inch brass strip, fiber washer, rubber feet, wooden knob, two 1½-inch brass "L"-brackets, assorted brass round-head wood screws.

Construction and Operation of a Simple Homemade Radio Receiving Outfit

The 1922 Bureau of Standards publication, *Construction and Operation of a Simple Homemade Radio Receiving Outfit* [1], is perhaps the best-known of a series of publications on radio intended for the general public at a time when the embryonic radio industry in the U.S. was undergoing exponential growth.

While there were a number of earlier experiments with radio broadcasts to the general public, most historians consider the late fall of 1920 to be the beginning of radio broadcasting for entertainment purposes. Pittsburgh, PA, station KDKA, owned by Westinghouse, received its license from the Department of Commerce just in time to broadcast the Harding-Cox presidential election returns. In today's world where instant global communications are commonplace, it is difficult to appreciate the excitement that this event generated.

News of the new development spread rapidly, and interest in radio soared. By the end of 1921, new broadcasting stations were springing up all over the country. Radios were selling faster than companies could manufacture them. The demand for information on this new technology was almost insatiable. The Radio Section of the Bureau of Standards provided measurement know-how to the burgeoning radio industry as well as general information on the new technology to the public. Letters to the Bureau seeking information on radio technology began as a trickle, and then soon became a flood. Answering them became a burden.

Circular 120, published in April 1922, began: "Frequent inquiries are received at the Bureau of Standards for information regarding the construction of a simple receiving set which any person can construct in the home from materials which can be easily secured. This publication has been prepared to meet these inquiries." The circular also noted that the Bureau was cooperating with the Bureau of Markets and Crop Estimates of the Department of Agriculture. In the early days of radio, broadcasts to farmers of crop price reports were an important public service.

In 1921 factory-made radios were very expensive. Many of them cost more than $2000 in today's dollars, and less affluent families could not afford to have one. When the Bureau published this circular (which sold for only five cents) on how to build a simple crystal radio at home, it was an instant success. Thousands of orders

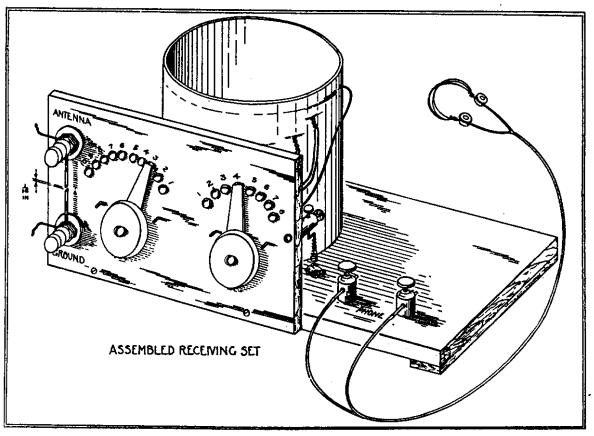


Fig 1. The crystal radio described in Circular 120.

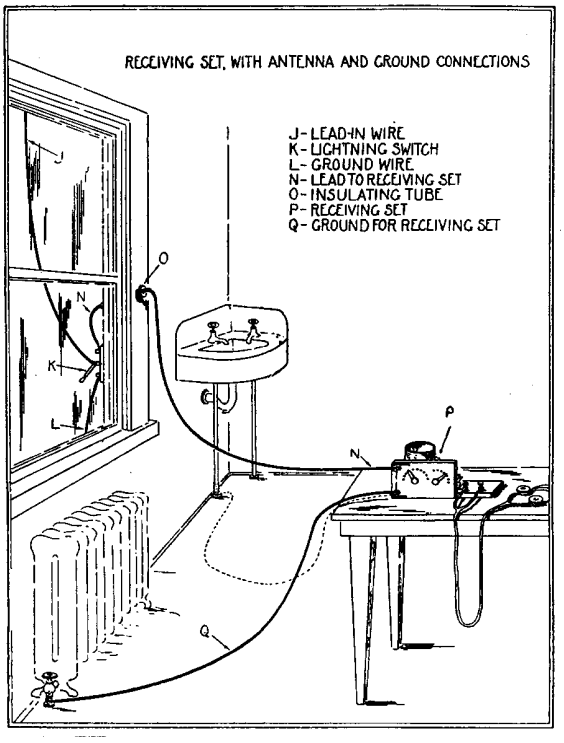


Fig. 2. Diagram showing the antenna and ground connections to the radio.

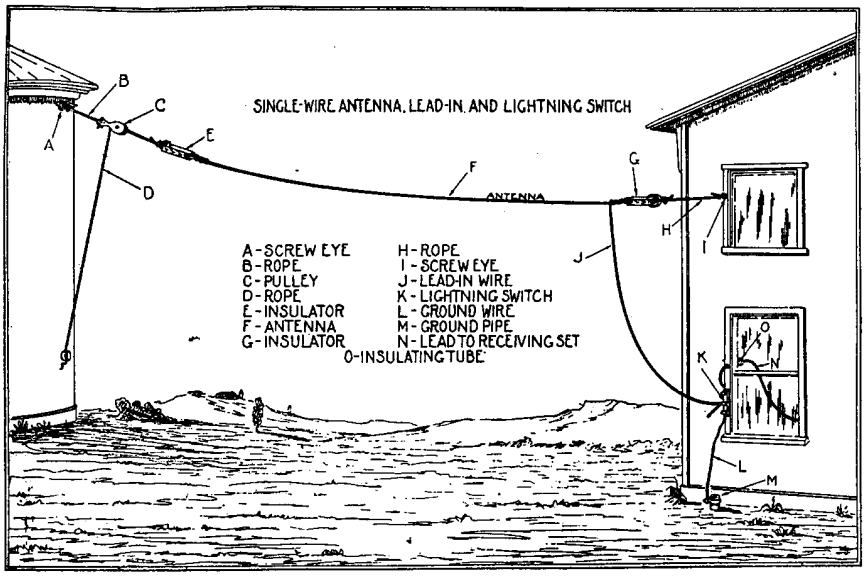


Fig. 3. Details of the outdoor antenna recommended for use with the crystal radio.

flowed in from every state. The circular showed how almost any family having a family member handy with simple tools could make a radio. The circular states "Satisfactory results have been obtained from sets constructed according to these instructions by persons having no previous experience with radio." The authors stated that the cost of the materials for the radio and associated headphones and antenna system should not exceed $10.

Newspapers around the country urged readers interested in radio to order a copy, and in many cases they reprinted the entire article. Even today, antique radio collectors searching attics and basements often come across dusty home-built crystal radio sets that were obviously constructed using the Bureau's plans. No doubt the remarkably rapid growth of the radio industry in the U.S. during the 1920s (analogous to the growth in Internet-related businesses today) was aided considerably by information provided to the public by the Bureau, such as this circular and the other items noted in the bibliography [2-6].

Circular 120 proved such a success that the Bureau published a number of other circulars on radio technology intended for the general public, as noted in the bibliography. All were popular. In some cases, commercial radio manufacturers copied the circuits proposed by the Bureau. During the same era, other radio publications of the Bureau captured public attention as well, though perhaps in smaller quantities than Circular 120. *Radio Instruments and Measurements* [7], Circular 74, published first in 1918 with a second edition published in 1924, became a best seller in the post-WW I era. Its principal authors were most likely J. H. Dellinger, J. M. Miller, F. W. Grover, and G.C. Southworth. The authors of Circular 120 are not stated, but some of those same individuals likely contributed to it. J. Howard Dellinger's story is particularly interesting. He came to the Bureau in 1907 as a laboratory assistant at $900 per year, and retired as Chief of the Central Radio Propagation Laboratory in 1948. He headed the Radio Section for much of its existence. Appendix D of Reference [9] is a biography of Dellinger.

The Principles Underlying Radio Communication [8], Signal Corps Radio Communication Pamphlet No. 40, published in 1919 and issued in a revised edition in 1922, is another interesting example of a popular radio book. Written by the Bureau of Standards for training Army Signal Corps officers, it was an expanded and updated version of Circular 74, noted above, but was also widely used by college students and others studying radio. In spite of the "pamphlet" designation, it was actually a 600-page book. Thomas Edison said ". . . This is the greatest book on this subject that I have ever read, and I want to congratulate you and your Bureau on its production."

While the general public was interested in radio information for the lay person, the fledgling radio industry needed in-depth technical assistance from the Bureau, which it got. The Bureau helped radio manufacturers make better measurements for quantities such as resistance, capacitance, voltage, and frequency. Results of Bureau research on antennas and radio wave propagation, direction finders, vacuum tube characteristics, and testing were eagerly utilized by major

companies such as General Electric, Westinghouse, Western Electric, and RCA, as well as by a myriad of small radio companies no longer in existence. At the same time that the Bureau was providing information to people constructing radios on their kitchen tables, Radio Section staff members were presenting technical papers on radio at conferences and in journals such as the *Proceedings of the Institute of Radio Engineers.* A good example of the latter is the classic work by John Miller, who discovered and explained "the Miller effect"—a feedback effect in triode vacuum tubes that limited their ability to amplify at radio frequencies [9].

Circular 120 described in detail not only the crystal radio itself but also how to construct an antenna and ground system. To minimize the cost, the circular suggested winding the tuning coil on a discarded cylindrical oatmeal box. For years afterwards, home experimenters used oatmeal boxes as coil forms for homemade radios. The crystal radio did not require batteries, but it did require the user to purchase a commercially made set of headphones (or telephone receivers as they were called in those days), since that accessory was not suitable for home construction.

One can just imagine the thrill of a farm family that had never heard a radio broadcast before trying the new set for the first time, and one of the family members crying out, "Come listen! I am hearing something!" The electronics revolution was underway and the Bureau played a significant part in accelerating the new technology.

Prepared by Brian Belanger.

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Teen Leads Space Station Contact, p. 88

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FEBRUARY 2014

CQ

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See page 8

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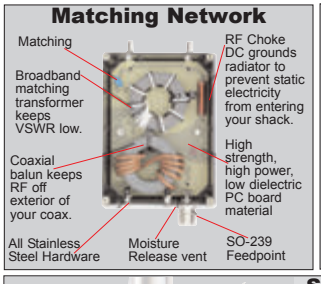

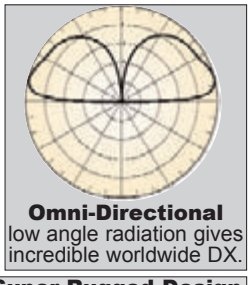

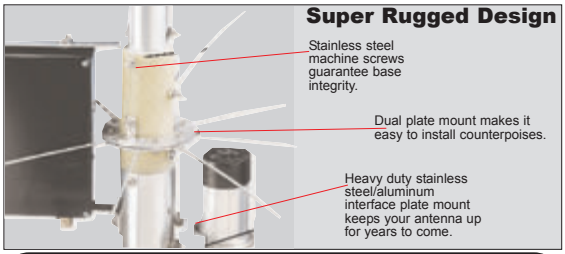

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Unrest in South Sudan Silences Ham Operators

The two licensed amateurs in South Sudan both report that they have taken their stations off the air and shipped their equipment out of the country "in order to avoid misunderstandings" with either side in the growing ethnic conflict that as of late December had claimed over a thousand lives. According to a report on the ARRL website, Massimo Stella, Z81B (also IZØEGB), and Diya al Asadi, Z81D (YI1DZ), said they expect to be off the air "for a while." Stella was in Europe while al Asadi remained in South Sudan. Both of them had been part of the Z81X operation from South Sudan in November, along with *CQ* DX Editor Wayne Mills, N7NG, who reports on that expedition and associated activities in his column this month (see page 94; note that Wayne filed his column prior to the outbreak of violence in the country).

Communications Act Update Planned

Citing the need for the law to keep pace with rapidly changing communications technology, the two Congressmen in charge of telecommunications policy in the House of Representatives say they are planning a comprehensive review and update of the Communications Act. Representatives Fred Upton (R-MI), Chairman of the House Energy and Commerce Committee, and Greg Walden, W7EQI (R-OR), Chairman of the Communications and Technology Subcommittee, say the effort will take over a year. Walden said the Communications Act, first passed in 1934 and last updated in 1996, is now "painfully out of date" and said the committee's goal is "to make sure this critical sector of our economy thrives because of the laws around it, not in spite of them." No timeline was given in the committee's announcement.

Listen for W100AW

The ARRL has been authorized by the FCC to use the special callsign W100AW during 2014, the League's centennial year. According to the *ARRL Letter*, the special call will be used in addition to W1AW, depending on the circumstance. Contacts made from the Maxim Memorial Station at ARRL Headquarters will use W100AW, as will those from regional conventions and during the IARU HF Championship this summer. However, regular bulletins and code-practice transmissions will still use W1AW, as will stations operating in the "W1AW WAS" program from each of the 50 states.

Solar Scientists: Cycle 24 Weakest in a Century

The current solar cycle is the weakest of the past ten cycles, according to four solar scientists reporting on "space weather" at the fall meeting of the American Geophysical Union. The *ARRL Letter* reports that the four—Nat Gopalswamy and Martin Mlynczak of NASA, Stanford's Leif Svalgaard and Joe Giacolone of the University of Arizona—say a weak polar magnetic field during Cycle 23 is thought to be the mechanism behind the weakness of Cycle 24. By continuing to track the polar magnetic field, said Svalgaard, it should be possible within a few years to predict the robustness (or lack of robustness) of Cycle 25. The group did not directly address the impact of the current weak cycle on radio propagation.

Hams Say "HI" to Juno Spacecraft

Hams around the world made coordinated transmissions last fall on 10 meters, sending "HI" in very slow Morse code to the sensors on the Juno spacecraft as it swung past Earth for a gravity boost on its journey to Jupiter. A video on the activity has been produced by NASA's Jet Propulsion Laboratory and is available on YouTube at <http://bit.ly/18PeGcI>. A more detailed article on the activity is planned for the March issue of *CQ Plus*.

Radio Arcala Tower Collapses in Storm

The massive 330-foot tower supporting the 160- and 80-meter Yagis at Radio Arcala's OH8X contest station in Finland collapsed under high winds in an early December storm that left some 200,000 families in the country without power. The 3-element 160-meter beam was reportedly the world's largest amateur radio antenna.

According to the *ARRL Letter*, group spokesman Jarmo Jaakola, OH2BN, speculated that the automation designed to enable the array "to find its most comfortable position in high winds" may have locked up and caused the structure to corkscrew. Jaakola said the tower—featured on the cover of the November 2011 cover of *CQ*—collapsed "peacefully" and caused no damage to any people or surrounding structures. At press time, inspectors were still trying to determine the exact cause of the failure.

Voice of Russia Leaves Shortwave

Many hams who came of age during the Cold War recall listening to the thinly disguised propaganda broadcasts of Radio Moscow booming in on their shortwave receivers. After the fall of the Soviet Union, the station was renamed the Voice of Russia but maintained a busy shortwave broadcast schedule. It began cutting back last year, though, and as of January 1, 2014, eliminated all shortwave broadcasting. The station will continue to reach out to international audiences via medium-wave broadcasting and the internet.

Astronaut-Hams Fix ISS Cooling System

Two American astronaut-hams repaired the International Space Station's cooling system during a dramatic Christmas Eve spacewalk, replacing a failed ammonia pump. According to *Newsline*, astronauts Mike Hopkins, KF5LJG, and Rick Mastracchio, KC5ZTE, each took two spacewalks as part of the repair mission, removing the faulty pump on December 21st and installing a new one on the 24th. The astronauts were outside the station for more than seven hours, in part because an ammonia fluid line sent frozen flakes of the substance right at them, with some sticking to their spacesuits. They had to wait until the ammonia on their suits dissipated before re-entering the station to avoid possible contamination.

In a related story, the *ARRL Letter* reports that former astronaut Susan Helms, KC7HNZ, has renewed her interest in amateur radio, regaining both her license and her callsign, which she'd allowed to expire in 2005. Helms, a Lieutenant General in the Air Force, is now the senior officer at Vandenberg Air Force Base in California. Her interest reportedly was reignited by an activity hosted by the Satellite Amateur Radio Club, W6AB, for Scouting's 2013 Jamboree on the Air. During a tour on the International Space Station in 2001, Helms used NA1SS for the first ARRL Field Day operation from space.

Ham Industry Changes

News of note from the ham radio industry:

• UK-based InnovAntennas has acquired Force 12 antennas of Bridgeport, Texas and consolidated its manufacturing activities with InnovAntennas America's facility in Grand Junction, Colorado. According to the InnovAntennas Facebook page, the combined facility "is in full swing and we are shipping, building and shipping antennas … daily."

• At the end of December, Japanese amplifier manufacturer Tokyo Hy-Power filed for bankruptcy and closed its manufacturing plant near Tokyo. The company's main website has also been closed down, although its US website remains active. According to a report on the ARRL website, company President and CEO Nobuki Wakabayashi, JA1DJW, said "the recent depression in the industrial RF power products area has led to the very difficult financial position."

• CQ Communications has realigned its publication lineup, merging *Popular Communications, CQ VHF* and *WorldRadio Online* into *CQ Plus*, a supplement to the digital edition of this magazine, effective with this issue. See "Zero Bias" on page 8 for more information.

Additional and updated news is available on the Ham Radio News page of the CQ *website at <http://www.cq-amateur-radio.com>. For breaking news stories, plus info on additional items of interest, sign up for* CQ*'s free online newsletter service. Just click on "CQ Newsletter" on the home page of our website.*

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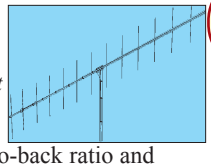

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CQ February 2014

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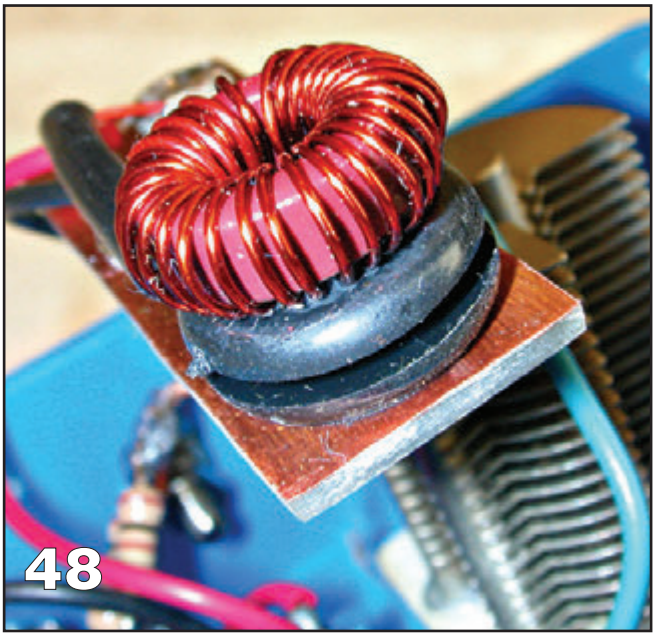

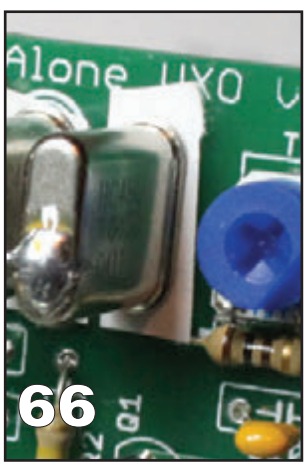

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Welcome to CQ Plus!

Beginning with this issue, *CQ*'s digital edition includes a supplement called *CQ Plus*. It includes content from *Popular Communications, CQ VHF* and *WorldRadio Online*, all of which have been merged into this new, expanded, version of *CQ*. Please note that this additional content is available only in the digital edition. There are no changes to the print edition. Print subscribers interested in adding a *CQ* digital subscription should contact our office for a special offer.

CQ plus *(digital edition only)*

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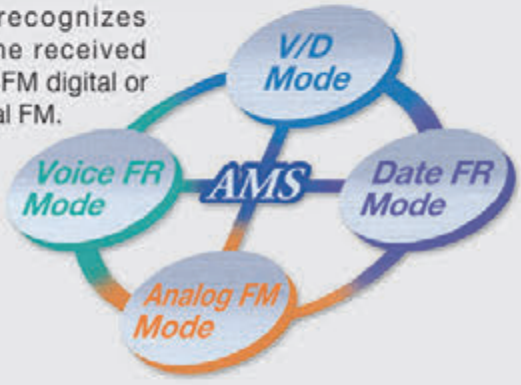

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zero bias – a CQ editorial

BY RICH MOSESON,* W2VU

CQ Plus – We're Changing to Better Serve You

How did you get started in the radio hobby? Did you dive right into ham radio, or did you start out doing something with an easier learning curve, such as shortwave listening or CB? Do you still enjoy some of those other aspects of hobby radio?

I certainly didn't start out in the deep end of the pool. I got started at about age 10, lying in bed late at night, listening to faraway AM broadcast stations in between the locals and trying to figure out where they were. This was followed by cold winter nights sitting in the family car, after I realized that the car radio was more sensitive than the old tube set in my bedroom (and had an external antenna outside the house, but I didn't know about that stuff yet). Then came CB, shortwave listening and finally, ham radio. I still dabble in SWLing and like to listen to airplanes and local public safety folks on my scanner.

I don't think my experience is at all uncommon. Most every ham I've ever met got started in a different branch of the radio hobby and worked his or her way up to the ranks of a licensed amateur. Do you still enjoy other parts of the radio hobby? I'm guessing yes. That's why virtually every HF transceiver on the market today includes a general coverage shortwave receiver, and why virtually every VHF/UHF transceiver includes wide coverage receive and scanning capabilities. We have collectively demanded it because we don't want to be limited by edges of the ham bands.

Very few of us are one-dimensional radio hobbyists. Yes, we may have a favorite part of the hobby, but most of us keep our fingers in other parts as well. Yet, hobby radio magazines have tended to be one-dimensional, specializing in one aspect of the hobby or another. Our focus in *CQ* has always been nearly exclusively on amateur radio, while our sister magazines have focused on other areas, either the broader listening part of the radio hobby or even narrower interests, such as contesting or VHF/UHF hamming. But no one magazine is really meeting all of your radio needs.

Well, we're about to change that. We're returning to the pre-1945 days of our predecessor magazine, *Radio*, and transforming *CQ* into your one-stop support center for all types of hobby radio!

Introducing *CQ Plus*

Beginning this month, selected content from our three sister publications, *Popular Communications, CQ VHF* and *WorldRadio Online*, will be incorporated into *CQ*'s digital edition as a supplement to be called "*CQ Plus.*" Take a look at our expanded table of contents (appearing in the print issue as well as the digital edition) to see what new offerings we have for you this month, from an update on the state of AM broadcasting and the FCC's efforts to "fix" it to an interview with a pirate broadcaster who regularly runs for President! The best of all four of our magazines is now combined into one—this one!

For our print readers, nothing will change. You will be getting the same *CQ* you've always gotten, with the same great features and columns you've come to expect. Digital subscribers will get both the print content and *CQ Plus* in a seamless, expanded, digital edition. Subscribers to *Pop'Comm, WRO* and *CQ VHF* have had their subscriptions transferred to *CQ*, with specifics based on the type of subscription they've had (print or digital) and the value of the remaining issues. Questions should be directed to our circulation department at <circulation@cq-amateur-radio.com> or 1-800-853-9797.

Yes, there is also an economic component at play here. First of all—as we've said here before—even though ham radio is growing at a nice pace, we're not seeing those increased licensing numbers translating into growth in radio clubs, ARRL membership or *CQ* readership. All of these are stable, but not keeping pace with growth in the number of licensees. Beyond amateur radio, the number of shortwave broadcast stations on the air is steadily shrinking, as cash-strapped governments decide they can effectively communicate with the rest of the world via the Internet and satellite, and without needing to maintain and operate expensive radio transmitter sites. Scanning options are shrinking as well, as more public safety agencies encrypt their (now mostly-digital) communications in the name of added security.

The bottom line for us is that *CQ* has been subsidizing *Pop'Comm, WRO* and *CQ VHF*. We can't continue to do that. Add in the glacial pace of the overall economic recovery and this has resulted in the delivery delays of both print and digital editions that have frustrated us as much as they have frustrated all of you. We are looking forward to a bigger, better, more comprehensive, *CQ* bringing us greater stability while continuing to serve our loyal readers throughout the radio hobby. We greatly appreciate your patience and your support as we have navigated these choppy waters, and we welcome the readers of *Pop'Comm, WRO* and *CQ VHF* to the pages of *CQ* and *CQ-Plus.* And speaking of which…

QRP Special

This month's issue is our annual **QRP Special**. We're focusing this year on the wide variety of ham radio activities that fall under the low-power umbrella, from operating QRP during lunch to using digital modes that copy signals you can't even hear! Of course, we've got non-QRP articles as well, from N2IRZ's column on how to build an inexpensive software-defined receiver to N2GA's fascinating suggestion that we try to make contesting a spectator sport as well as a participation sport.

A Sad Note

It is with great sadness that we pass along the news that former *CQ* Editor and CQ Amateur Radio Hall of Fame member Arne Trossman, W4EIP (ex-W2DTJ), became a Silent Key in November. In addition to his time at *CQ* (1960–64), Arne helped develop the Cosmophone, an early amateur SSB transceiver.

On a personal level, Arne was one of my first ham radio "Elmers" or mentors. I grew up a couple of towns away from Arne and he was my Radio Merit Badge counselor when I was a Boy Scout. I still remember going down to his basement ham shack and workbench, where he taught me to solder and helped me learn Morse code. It wasn't until many years later that we made the connection, realizing it was his work with a kid trying to earn a merit badge that set me on the path to eventually become his successor in the *CQ* editor's chair. 73, Tross. We'll miss you.

73, Rich, W2VU

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announcements

RICHMOND, VIRGINIA – Richmond Amateur Telecommunications Society (RATS) Frostfest 2014, Saturday, February 1 at the Richmond Raceway Complex. Contact: RATS/Frostfest, P.O. Box 14828, Richmond, VA 23221-0828. Commercial booth sales: Mike Hacket, AC5PT, (804) 657-7038. Website: <http://www.frostfest.com>.

ELYRIA, OHIO – Northern Ohio Amateur Radio Society Winter Hamfest & Computer Show 2014, Sunday, February 2 at the VFW Post 1079. Contact: Darlene, KA8VTS, (216) 398-8858. E-mail: <dohman@roadrunner.com>. Website: <http://www.noars.net>. (Talk-in 146.70– [PL 110.9])

WASHINGTON, PENNSYLVANIA – Washington Amateur Communications special event station W3C from 1400 to 1600 UTC daily, Wednesday, February 5 through Sunday, February 10. For a complete list of frequencies and photos visit <http://www.wacomarc.org>. Contact: Bill Steffey, NY9H, (224) 436-4262. E-mail: <ny9h@arrl.net>.

BRIGHTON, COLORADO – Aurora Repeater Association, The Cherry Creek Young Amateur Radio Club, and Rocky Mountain Ham Radio Swapfest, Sunday, February 9 at the Adams County Fairgrounds. Contact: ARA, P.O. Box 471802, Aurora, CO 80047-1802. Wayne Heinen, NØPOH, (303) 699-6335. E-mail: <n0poh@arrl.net>. Website: <http://n0ara.org>.

YUMA, ARIZONA – Yuma Amateur Radio Hamfest Organization and the Amateur Radio Council of Arizona 10th Annual Yuma Hamfest, Friday, February 14 and Saturday, February 15 at the Yuma County Fairgrounds. Contact: YARHO, P.O. Box 1843, Yuma, AZ 85366-1843. E-mail: <info@yumahamfest.org>. Website: <http://www.yumahamfest.org>. (Talk-in 146.840– [PL 88.5])

B**ROOKFIELD, WISCONSIN – Milwaukee Radio Amateur's Club and the Milwaukee Area Amateur Radio Society Mid-Winter Interclub Swapfest**, Saturday, February 15 at the Channel 10 Auction Building. Contact: MRAC, P.O. Box 26233, Milwaukee, WI 53226. Phone: (414) 459-9741. E-mail: <w9rh@arrl.net>. Website: <http://www.w9rh.org>.

GLOUCHESTER CITY, NEW JERSEY – Glouchester City Amateur Radio Club Hamfest, Saturday, February 15 at the Pine Grove Fire Association Hall. Website: <http://www.nj2gc.org>. (Talk-in 146.820 [PL 131.8] or 447.775 [PL 146.2].

OAK HILL, WEST VIRGINIA – Plateau Amateur Radio Association Hamfest, Saturday, February 15 at the Lewis Community Center. Contact: PARA Hamfest, P.O. Box 96, Fayetteville, WV 25840. General contact: Jane Hardy, WV8JH, (304) 640-1120. E-mail: <wv8jh@frontier.com>. Charlie Hardy, WV8CH, (304) 719-2241. E-mail: <wv8ch@arrl.net>. (Talk-in 146.790 [PL 100])

SALEM, OREGON – **Salem Repeater Association Salem Hamfair**, Saturday, February 15 at the Polk County Fairgrounds. Contact: Wayne Silver, KE7ANM, (971) 599-1270. E-mail: <hamfair@w7sra.org>. Website: <http://www.w7sra.org>. (Talk-in 145.33– [PL 186.2])

LIVONIA, MICHIGAN – Livonia Amateur Radio Club Swap-N-Shop, Sunday, February 16 at the Civic Park Senior Center. Contact: Livonia ARC, P.O. Box 51532, Livonia MI 48151-0532. Phone: (734) 941-5043. E-mail: <k8uns@arrl.net>. Website: <http://www.livoniaarc.com>. (Talk-in 145.35 [PL 100] or 146.52)

MANSFIELD, OHIO – Intercity Amateur Radio Club, Inc. MidWinter Hamfest & Computer Show, Sunday, February 16 at the Richland County Fairgrounds. Contact: Danny Bailey, KB8STK, 70 Euclid Street, Shiloh, OH 44878. Phone: (419) 896-3603 (after 3 p.m.). Website: <http://www.w8we.org>. (Talk-in 146.94 [PL 71.9])

NEW PROVIDENCE, NEW JERSEY – New Providence Amateur Radio Club Annual Auction, Friday, February 21 at the New Providence High School. Contact: Barry, K2JV, <bgcohenusa@verizon.net>. Website: <http://www.nparc.org>. (Talk-in 147.225+ [PL 141.3])

BISMARK, NORTH DAKOTA – Central Dakota Amateur Radio Club Hamfest, Saturday, February 22 at the St. Mary's Grade School. Contact: Dick Veal, KAØETO, (701) 223-7481. E-mail: <georgev@bis.midco.net>.

BROWNSBURG, INDIANA – Hendricks County Amateur Radios Society Brownsburg Hamfest, Saturday, February 22 at the American Legion Post 331. E-mail: <hcars46122@gmail.com>. (Talk-in 147.015+)

DALTON, GEORGIA – Dalton Amateur Radio Club Inc. Dalton Hamfest, Saturday, February 22 at the North Georgia Fairgrounds. Contact: DARCI, P.O. Box 211, Dalton, GA 30722-0211. Vendor contact: James Jordan, K4FLG, (706) 278-0630 or David Stanley, WI4L, (706) 537-5090. Website: <http://www.w4drc.com>. (Talk-in 145.230–)

GEORGE, WASHINGTON – Columbia Basin DX Club, WS7G, special event station from 0001 UTC, Saturday, February 22 to 0001 UTC, Sunday, February 23. Frequencies include 14.250, 18.135, and 3.850 ±QRM. QSL via W7BJN. Website: <http://cbn.homestead.com/WS7G.html>.

LaPORTE, INDIANA – LaPorte County Amateur Radio Club Cabin Fever Hamfest, Saturday, February 22 at the LaPorte Civic Auditorium. Contact: LPCARC, P.O. Box 148, Michigan City, IN 46361. E-mail: <cabinfeverhamfest@gmail.com>. Website: <http://lpcarc.org/hamfest>. (Talk-in 146.610 [PL 131.8])

PERRY, IOWA – Hiawatha Amateur Radio Club WinteRFest, Saturday, February 22 at the Crossroads Church. Contact: Bob Dittert, (515) 465-2383. E-mail: <n0qix@arrl.net>. Vendor e-mail: <hamfest@harciowa.org>. Website: <http://www.harciowa.org>. (Talk-in 145.190 [PL 114.8])

SOUTH BURLINGTON, VERMONT – Radio Amateurs of Northern Vermont HAM-CON, Saturday, February 22 at the Holiday Inn Convention Center. **Special event station, W1V** will be on the air during HAM-CON. Contact: RANV, (802) 879-6589 afternoon and evenings. Email: <w1sj@arrl.net>. Website: <http://www.ranv.org>. (Talk-in 145.15– or 146.67)

ANNANDALE, VIRGINIA – Vienna Wireless Society Winterfest,The National Capital Area Ham Radio Fair, Sunday, February 23 at the Northern Virginia Community College-Annandale Campus. Contact: Vienna Wireless Society, P.O. Box 418, Vienna, VA 22183. General contact: Jack, AI4SV, (314) 266-8426. Email: <winterfest2014@viennawireless.org>. Vendor contact: Doug, AK4AO, (703) 698-6158. E-mail: <tablesales2014@viennawireless.org>. Website: <http:// www.viennawireless.org>.

CASTLE SHANNON, PENNSYLVANIA - Wireless Association of South Hills WASHfest 2014, Sunday, February 23 at the Castle Shannon VFW Memorial Hall. Contact: Carol Danko, KB3GMN, (412) 884-1466. E-mail: <n3sbf@comcast.net> or <washarc@yahoo.com>. Website: <http://www.n3sh.org>. (Talk-in 146.995– or 443.650+ [PL 131.8])

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Welcome Once Again to the Fun World of QRP Operating!

BY CAM HARTFORD,* N6GA

Hello, and welcome to *CQ*'s third annual QRP Special edition! The first QRP Special was published in April 2012, the second in March 2013, and this year we find ourselves in February 2014. Management at *CQ* Central deemed the forward creep of one month a year a necessity to accommodate scheduling changes in contest reporting . I suspect that it's something entirely different—that it's really the pressure of the QRP world wanting to make itself known. As I noted two years ago, there are QRP contesters, DXers, SSBers, digi-ops, home-

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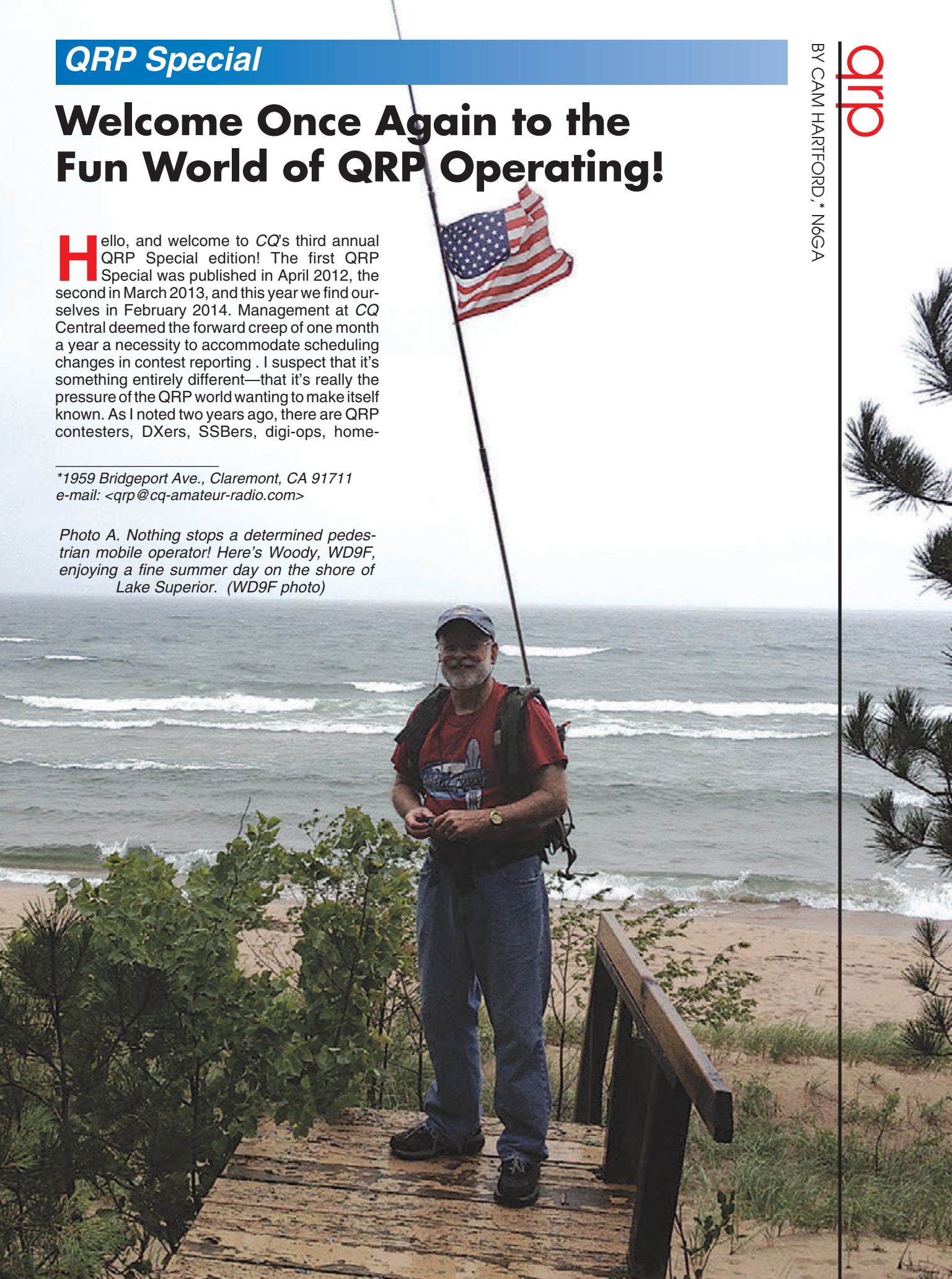

Photo A. Nothing stops a determined pedestrian mobile operator! Here's Woody, WD9F, enjoying a fine summer day on the shore of Lake Superior. (WD9F photo)

brewers, kit-builders, to-the-fielders, SOTAers, park-bench operators, fox chasers, net QNIers, and ragchewers. Oh yes, don't forget the pedestrian mobile ops. Read on!

Walkin' and Talkin'

I recently received this report from Woody Hester, WD9F (photo A). It seems Woody was bitten by the pedestrian mobile bug, and after an initial shakedown cruise has been stepping out in fine P/M style. Here's Woody's account:

Photo B. WD9F's p/m rig. (WD9F photo)

"After hearing Ed Breneiser's (WA3WSJ) excellent presentation on QRP pedestrian mobile operation at Four Days in May 2012, and having worked the amazing 'Father of /pm operation,' Paul Signorelli, WØRW, operating pedestrian mobile in three different states (Colorado, Florida, and Texas) over the last few years, I just had to give this QRP/pm thing a go.

"I bought a like new ALICE Pack frame on eBay, an antenna mirror mount at a truck stop near my home, and then began attaching the needed items to the pack frame to make up my /pm station (see photo B). What did we ever do without bungie cords?"

Fortunately, Woody has graduated to a hybrid mounting system which incorporates both bungie cords and zip ties. I'm not a big fan of bungies, having seen too many accidents where bungies sprung when it was least convenient, but at least in this case there's nothing more at risk than a few radio bits.

Woody continues: "I ended up using an MFJ 9020 transceiver for 5 watts on 20 meters, a 5-amp/hour SLA battery, an LDG auto tuner, and my Palm mini paddles. For my antenna, I attached a Hustler mobile antenna with a 20-meter resonator and included a 14-foot drag wire (24-gauge black silky from The Wireman). Total height is right at 12 feet.

"My first outing started in my Central Illinois backyard. ... I then proceeded out onto the nearby Sangamon Valley Bike Trail, and in about an hour of walking, I netted five solid QSOs. Tom, KM4CU, in Jax, Florida gave me a 559; Jean, VE2JCW, near Montreal gave me a 579; John, KN5L, in Austin, TX a 539; Michel, VE2TH, in Quebec running a new KX3 a 559; and Christian,VE1/F8BMG, running an IOTA station in the St. Lawrence a 579."

Wow, nice run of contacts for a first outing! It's hard to believe that a person can reach that diverse a crowd with such a small antenna and low power, but as Dave Ingram, K4TWJ (SK), used to say, "QRP Rocks!"

Here's a little more detail on Woody's /pm setup: "It's all built on an "ALICE Pack" frame. 'Alice' stands for 'all-purpose lightweight individual carrying equipment'. It's the pack frame that was adopted by the U.S. Army as the new standard in 1972 or so. It's made out of aluminum, it's strong, and it really is light weight. The pack frame can be had on eBay almost any day for around $20 plus shipping. I also bought two optional aluminum shelves specifically designed for the ALICE frame. They

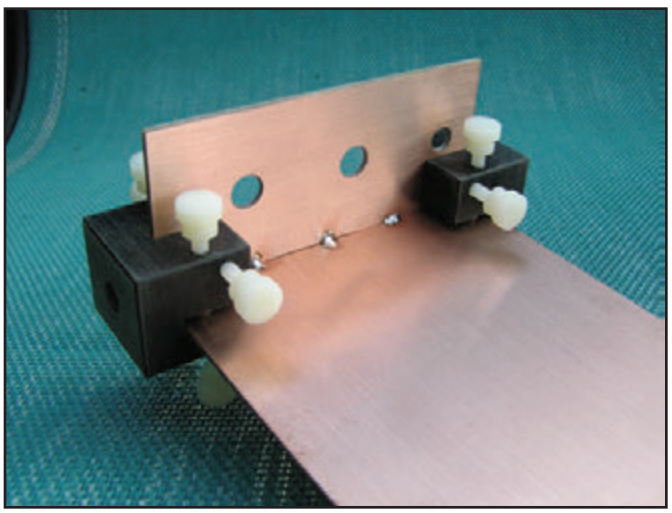

Photo C. A pair of SBSSs from QRPme hold this front panel in place while it's being soldered to the cabinet bottom.

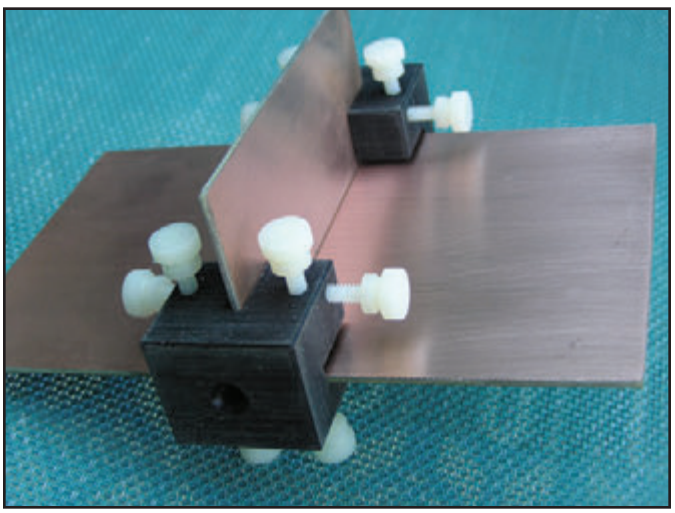

Photo D. Here the SBSSs are holding an interior shield in place for soldering.

were about $10 each on eBay and they really make it easy to attach things. ... Things I've learned on the trail are that I can always make contact with someone. Getting up high helps, too. We have a small hill (no mountains here in the flatlands) about a mile away, and when I get up on that it helps. The noise level drops way down when I get away from civilization and I hear very weak signals I don't normally hear. People watch me closely. Some ask what I'm doing, but most don't. They just give me a wide berth and keep moving. I always smile real big and give them a thumbs up. I do have to watch for overhead objects/tree branches, I carry a couple of extra drag lines in case one gets pulled off and I don't notice, and my earbuds and palm mini paddles are just perfect for this application."

This past summer Woody and his wife traveled to the southern shore of Lake Superior. "Yep, I took my pedestrian mobile gear along and spent several enjoyable hours hiking along the beach. The included picture was taken on a cold and windy day with light rain blowing horizontally down the beach. I still made several contacts. Highlights of my trip were two /pm to /pm contacts. One with Paul Signorelli, WØRW, who was /pm in Colorado; and Ed Breneiser, WA3WSJ, another widely acclaimed and accomplished /pm op who was pedestrian mobile on Pikes Peak. Ed had gone to Colorado from his home in Reading, PA to see the sights, operate pedestrian mobile, and visit Paul Signorelli. ... These contacts weren't planned. They took place on different days and just happened. I was thrilled. Paul had a great 589 signal from me, gave me a 559, and Ed and I were 559 to each other, all on 20 meters. That's my workhorse band for /pm. It's always open to somewhere. There's always someone around the QRP/SOTA watering holes, and my /pm antenna really likes it. I also worked Steve Galchutt, WGØAT, of Rooster and Peanut fame. Steve was at home in Colorado trying out his brand new ATS 4."

It sounds like Woody is now a true convert to world of QRP/pedestrian mobile. If you hear him around 14060, be sure to give him a call. He's probably out on the trail, ducking low-hanging trees and making QRP Qs!

Passing the Torch in the Northeast

A visit to the Small Wonder Labs website <http://www.smallwonderlabs.com/> yields a farewell message from Dave Benson, K1SWL, proprietor. It's not really a surprise, as Dave has been busy building his retirement home in the woods of New Hampshire while slowly reducing his involvement in kit production. Over the past 20 or so years the QRP community has benefited greatly from Dave's offerings. In this shack alone I can lay my hands on several SWL kits, including an early 2-board 20-meter transceiver that appeared in the *ARCI QRP Quarterly* from January 1993, an NN1G 20-20, a 20-meter DSW, a Warbler, a White Mountain SSB rig, and a couple of Rockmites. I even

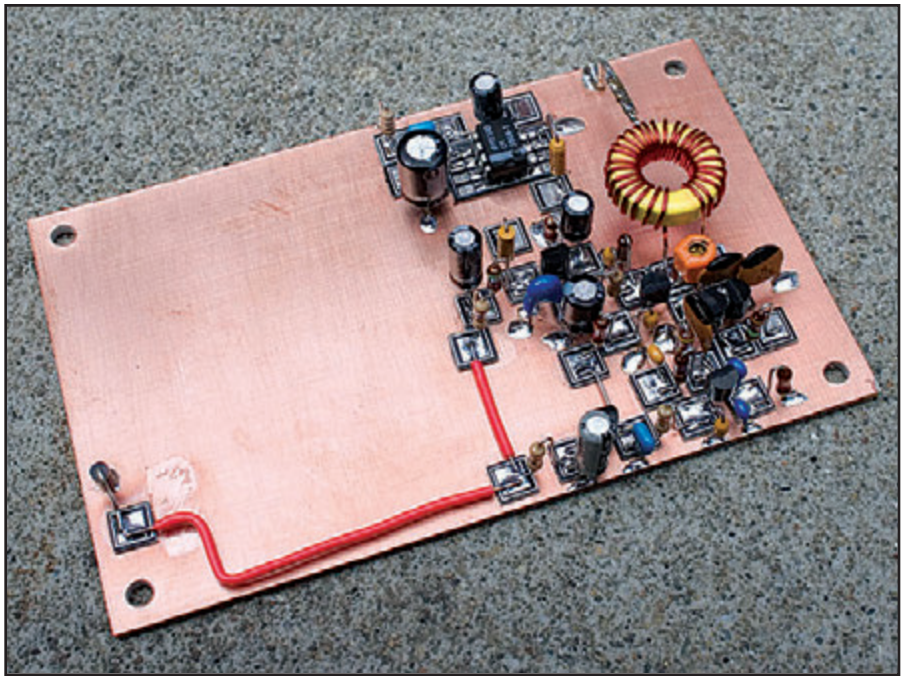

Photo E. Dave, AA7EE, used both MePADs and MeSQUAREs in this Manhattan project. Along the top edge of the board is an LM-386 soldered to a MePAD designed for an 8-pin IC. (AA7EE photo)

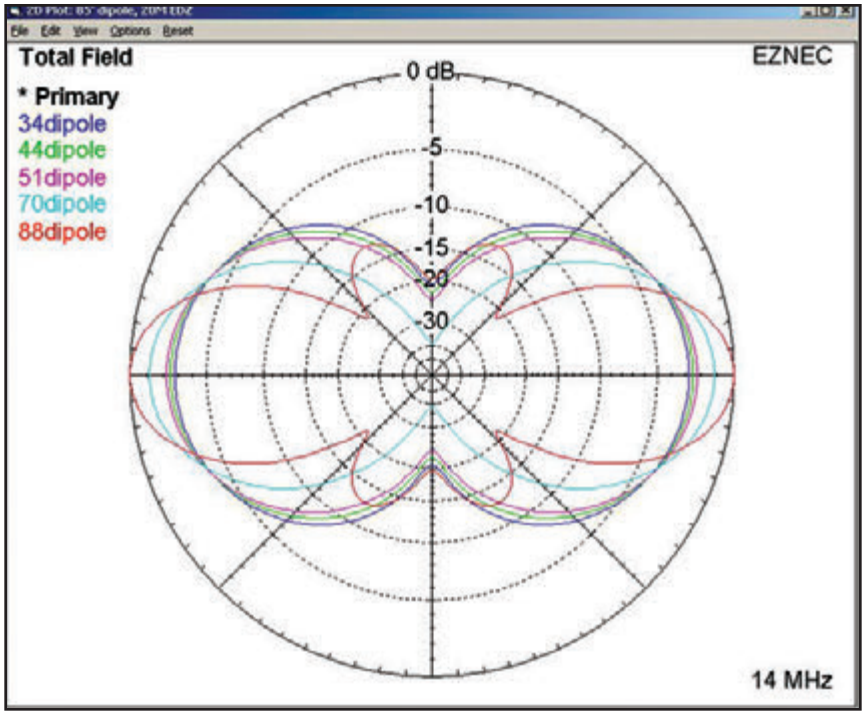

Fig. 1. EZNEC plots of several lengths of dipoles, all at 60 feet elevation and driven at 14 MHz.

have an 80-meter transceiver that I built from scratch from an article Dave published in the October 1991 issue of the *QRP Quarterly*. Thus, I hate to see Dave go. We all benefited greatly from his kits and support over the years, but he deserves a well-earned rest.

But wait! A little bit of Small Wonder Labs lives on! Rex Harper, W1REX, of QRPme (http://www.qrpme.com), the Tuna Tin place, has picked up the RockMite design and will continue to sell the very popular little rig. According to Rex: "The Small Wonder Labs RockMite design grew out of K1SWL's desire to offer a one-evening CW transceiver kit that would be both affordable and easy to construct. ... It appeared in *QST* magazine in 2003, and since then, it's been wildly successful. Approximately 8000 of these have been fielded so far!" Rex and Chuck Carpenter, W5USJ, have done some updating to the design, including bringing it up to current FCC requirements for harmonic suppression.

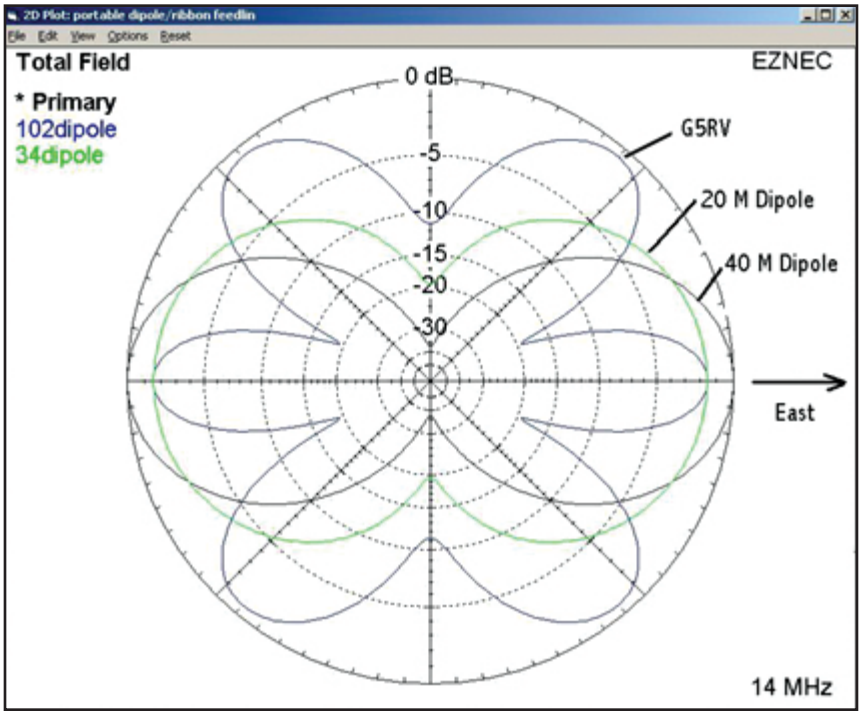

Fig. 2. Plots of a G5RV, 20- and 40-meter dipoles. Notice the difference between the lobes of the G5RV and the 40-meter dipole in the direction of interest.

More QRPme Bits

I was wandering around Rex's website and came across something I'd never seen before. They are called the "SBSS" which stands for—oh, I don't really know—maybe Solder Bot Square Something. Anyway, I saw these and immediate administered myself a thump on the forehead. Brilliant idea! Why didn't I think of that? There are lots of reasons which we won't waste time discussing here. Photo C shows a pair of the little things holding together a couple of pieces of PC board material ready for soldering. They hold the pieces at right angles so you can concentrate on doing the soldering ,and, hopefully, your chances of coming up with a straight, square enclosure for your next project will be greatly enhanced. You can also use them for making Tee joints, as in photo D, if you have the need to place a shield somewhere inside the box.

Also worth checking out are Rex's MePADS and MeSQUARES. (See photo E.) These little bits are made for the homebrew enthusiasts who favor the Manhattan style of construction. The sheet of Squares snap apart into hundreds of pre-tinned pads that you only have to glue down. Also, each Pads sheet consists of 28 pre-etched IC pads that you can snap apart and glue down. As Rex says, "They make it 'wicked easy' to mount both DIP and SOIC integrated circuits in your project." Photo E is an in-process shot taken by Dave Edwards, AA7EE, of a regenerative receiver he was making with the use of MePADS and MeSQUARES. Check out <http://aa7ee.wordpress.com/2011/07/24/the-wbr-a-simple-high-performance-regen-receiver-for-40m-by-n1byt> for an in-depth look at this homebrew project!

A Tale of Two Dipoles

Last September a couple teams of QRPers assembled with the intent of competing with the big boys in the CW Ops Open. The CW Ops run a couple of short sprints each month as well as an annual competition called the Open. You can find all the info about this organization at <http://www.cwops.org/>. Hank Greeb, N8XX, rallied the troops for this venture, and I have to say, fielding a team of Davids to take on the Goliaths is a bunch of fun. We never hurl a big enough rock to do serious damage, but we do place well enough to show what QRP can do.

What distinguishes this contest from many others is that there is an effort to include CW ops from around the world. Total operating time is 12 hours, but that is broken up into three 4-hour sessions timed for good propagation in Europe, Asia, and the Americas. For those of us in the Pacific time zone that means session One is from 5 to 9 PM Friday evening, session Two is from 5 to 9 AM Saturday morning, and session Three is from 1 to 5 PM Saturday afternoon. Taking into account the oddball session times and the fact that some local noise source had been making life very miserable at my home QTH, I made the decision to flee to the local mountains and do a portable setup. There might be fewer honey-do's up there, too.

This would be like a mini Field Day but without the crack team of experienced antenna erectors. I'd have to do this on my own, limiting me to fairly simple antennas. A 120-foot tower with stacked Yagis would be an excellent choice, but not this time. It was looking like I'd have to fall back on that dependable workhorse of antennas, the dipole.

From southern California most of the country is centered around an azimuth of 75 degrees, Europe is about 30 degrees and JA is at about 320. However, the sunspots had been puny in the days leading up to this contest and the outlook for DX was rather dim so I chose to optimize the antennas for in-country use. I could put up two, one aimed across the country and another aimed north-south so as to pick up the western states.

The next question facing me was exactly what dipole to use. Twenty meters would be the bread-and-butter band, but after dark 40 and even 80 came into play. There also could be some action on 15 meters, complicating the decision.

I lined up the usual suspects, which

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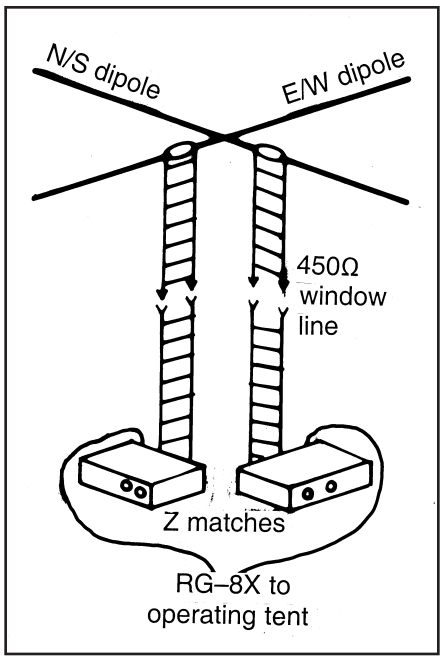

Fig. 3. Layout of the David vs. Goliath antenna farm.

were, from smallest to largest, 34 feet, 44 feet, 51 feet, 70 feet, 88 feet, and 102 feet. The smallest one is a 20-meter dipole, which would work fine for 20 and 15 but not for the lower bands. The 44 is known as a "Half 88." I believe L. B. Cebik popularized these sizes as usable for multi-band dipoles when fed with ladder line. The 51-footer is known as a "Half G5RV" for obvious reasons. The 70-foot antenna is a 40-meter half-wave dipole. The 88 foot antenna is, coincidentally, an Extended Double Zepp on 20 meters.

Fig. 1 is a composite of the radiation patterns of each one of these antennas on 14 MHz, all at a height of 60 feet. I've left off the G5RV for the sake of clarity. The 34-foot antenna shows the classic pattern of a half-wave dipole. As we lengthen the wire, two things happen: The gain in the lobe increases, while the width of the lobe, or its beamwidth, decreases. From the 34- to the 44- to the 51-foot dipole you can see a slight lengthening of the lobe (increasing gain), while at the same time the beamwidth narrows slightly. There's not a lot to choose from amongst these three antennas, as the gains are all within 1 dB of each other. There's a noticeable jump when we go to the 40-meter dipole, however, at 70 feet in length. Here the gain over the 20-meter dipole is about 1.6 dB and the beamwidth is noticeably narrower.

Referring back to fig. 1, the 88-foot antenna also shows a substantial jump in the gain, but once again at the expense of beamwidth. The Extended Double Zepp is a significant length for a dipole, because at 1.25 wavelengths, it's the longest dipole that exhibits its maximum directivity broadside to the wire. Any longer and the main lobes start to shrink and the little side lobes grow, spreading your power in many points around the compass.

The G5RV, at 102 feet, would have been a good antenna for 40 and 80 meters, but on 20 meters the main broadside lobe has shrunk to a skinny little thing. See the plots in fig. 2. This shows the G5RV, 20- and 40-meter dipole plots. On 20 meters, there's lots more signal headed across the country using the 40 meter dipole. However, I noticed something interesting about this antenna. The two biggest lobes are at about 40 degrees either side of the axis of the wire, meaning that if I needed to head for the hills for a DX contest I could conceivably put up a G5RV with the wire on a north-south axis and the big lobes pointed at Europe and Asia. Food for thought.

In the end, I chose to use the 40-meter dipoles. At 70 feet long this antenna is resonant on 40 meters and has some gain and directivity on 20 meters. The EDZ would have had more gain but at the expense of beamwidth. Additionally, on 15 meters at 88 feet of length the major lobes split to form an almost

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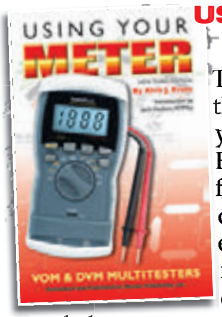

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perfect clover leaf, aiming a nice deep null across the country. It's never helpful to have your antenna aimed at every place but where the potential contacts are! Even the 40-meter dipole scatters the 15-meter signal around, but with two antennas at right angles I was able to get a lobe aimed sufficiently well to snag plenty of 15-meter Qs.

Having the two dipoles set at right angles covered the country pretty well. The patterns of the antennas overlapped somewhere in the direction of the Dakotas on both 40 and 20 meters. Switching between the two with the A/B antenna switch confirmed this. Mother Nature was thoughtful with her tree placement, so I was able to get the antennas up within about two degrees of perpendicular. This is necessary to minimize coupling between the antennas and hence skewing of the patterns. I fed both antennas with 450-ohm window line down to ground level, where I had a pair of homebrew Z-match tuners, and from there ran coax into the operating tent. Here the two coaxes went into the A/B antenna switch.

With this setup I had 40, 20, and 15 covered pretty well, but not 80 meters, and 80 would provide plenty of Qs if there were an easy way to incorporate it. The problem with running 80 meters on a 40-meter dipole is one of efficiency. The 40-meter dipole will radiate a signal on 80 meters, but the feedpoint impedance is highly reactive with a low resistive component. This will result in substantial feedline loss, even if you are using ladder line. Recommended reading on the subject is this page from the W8JI website at <http://www.w8ji.com/short_dipoles_and_problems.htm>. Modeling the 40-meter dipole on 80 meters in EZNEC shows a feedpoint impedance of 14.6 – j1110 ohms. Using the VK1OD transmission line loss calculator referenced by W8JI <http://vk1od.net/calc/tl/tllc.php>, we find the loss on the line to be 5.5 dB. Add to that another dB or so of loss in the tuner and the 5 watts coming out of your transmitter shrinks to 1 watt by the time it reaches your antenna.

What to do? I cut two pieces of wire 20 feet long. On each one I soldered a big alligator clip to one end and tied a 10-ounce fishing sinker on the other. When it came time on Friday night to change bands to 80 meters, I dropped one of the dipoles and clipped one of the wires to each end; the resulting antenna looking like an upside-down U. The resulting antenna was 110 feet long, so the feedline loss was substantially less. Something closer to four watts was actually making it to the feedpoint. Unfortunately, I hadn't visualized making this change after dark so things didn't work exactly as planned. When I hauled the antenna back up after clipping on the end wires, one end got tangled in an antenna-eating tree and all operations stopped for a good half hour of yanking and swearing by the puny beam of a flashlight. I got back on the air with 7 minutes left in the session and had a total of three 80-meter QSOs to show for it. *Note to self:* Visualize things better!

There were lots of lessons learned in this venture. The 40-meter dipoles worked well for the bands and the directions I wanted to cover. The 80-meter extensions would have served well but for what we'll call "cockpit problems." Most importantly, a high dipole can be a very effective antenna and shouldn't be overlooked if you are going out on a Goliath-slaying venture! (See fig. 3.)

Sign-Off

Thanks for riding along on this QRP adventure. Fun as it may be, nothing beats the real thrill of getting on the air and doing it yourself! 72/73, Cam N6GA

No time to get on the air? Consider W2LJ's approach to squeezing some quality QRP DXing into his busy day.

QRP à la Carte

BY LARRY MAKOSKI,* W2LJ

For those of us still actively participating in the workforce, it's a common tale of woe: Between full-time employment (in some cases, more than one job), kids' activities, civic or church responsibilities, home improvement tasks, house chores, etc., what time is left for amateur radio? By the time I get home from work, eat dinner, walk the dog, and do whatever other chore I have in front of me, I'm ready for bed ... not the shack! Weekends can be even worse with the myriad of home-improvement projects and yardwork tasks that seem to never end. What's an avid QRPer to do?

Fortunately, I have an hour thrust upon me each day which provides a little "off" time. That hour, otherwise known as lunchtime, has become a golden opportunity for getting in some QRP time. Instead of sitting in the cafeteria, or at my desk with a sandwich while absentmindedly gazing at YouTube, you will find me on the air, either scanning the bands or pounding out a CQ.

Quick Setup and Teardown

The key to successful lunchtime QRP operation is quick setup, easy-to-operate equipment, and quick disassembly when you're done. You don't want to spend your time fussing with equipment while you could be making contacts. If your setup is a permanently mounted mobile HF rig, that's all the better. Hopefully, you can turn your power output down to 5 watts and take off running. However, if you just tote along some QRP gear in a backpack like I do, then there are some considerations that must be made.

The first major consideration: From where will you operate? Your options may or may not be limited. If you're fortunate enough to work very close to a public park, this can be a perfect place to set up (photo A). At my last job, I had a park only five minutes away. I was able to get there, set up, and operate within a very reasonable amount of time (photo B). Others are not so lucky. In this case, you may want to consider just going out to your car in the company parking lot and setting up from there (photo C). Your circumstances will dictate your setup.

Antenna Choices

The next consideration is what to use for an antenna. While it has to be a com-

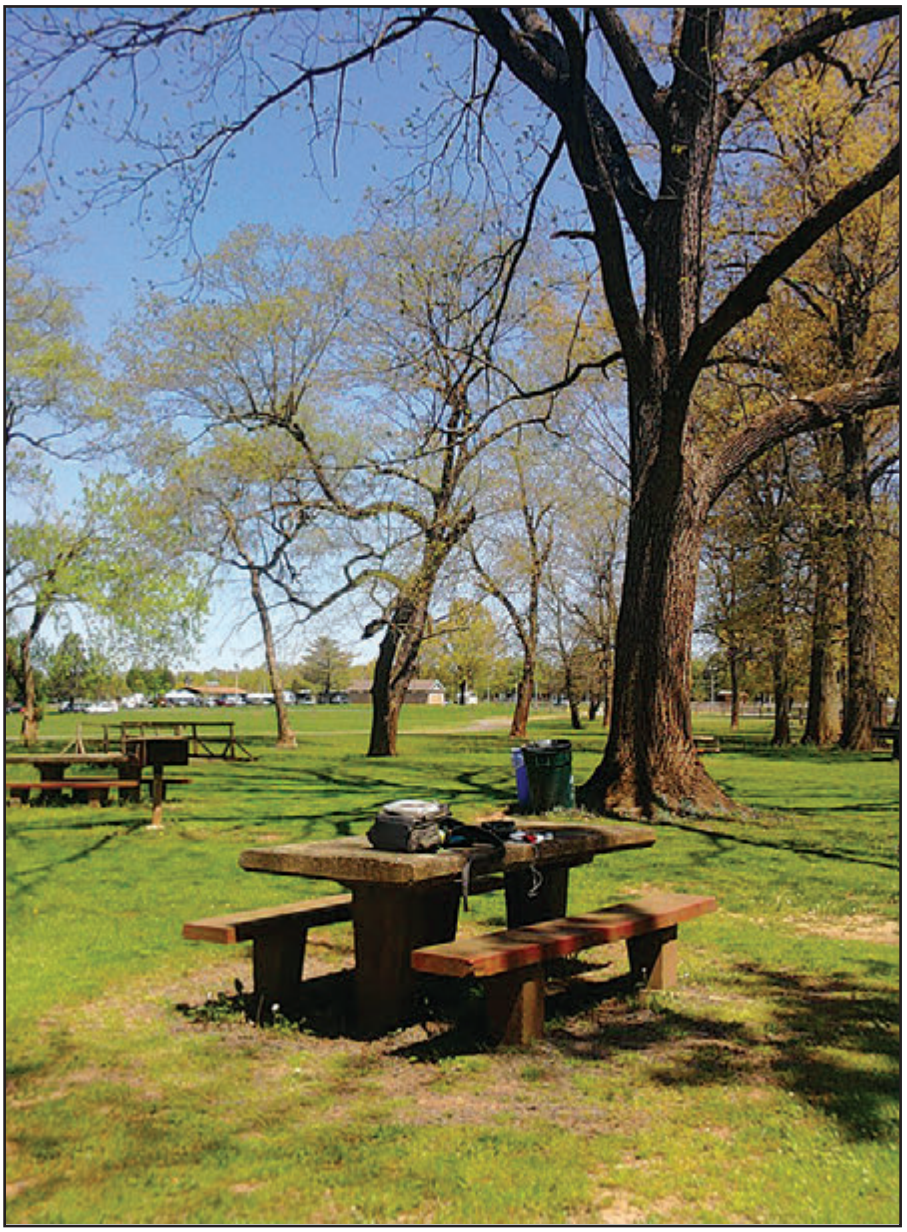

Photo A. A picnic table can make an ideal lunchtime operating position, if you're lucky enough to work near a public park. (Photos by the author)

*e-mail: <w2ljqrp@gmail.com>

Photo B. No picnic table? Try operating from your car. In this photo, we see a quick-setup antenna supported by a mast with a baseplate that sits underneath a tire.

promise for quick setup for a short portable operation, you do want it to be somewhat efficient. There's no fun in trying to make contacts while sending your 5 watts into a dummy load. Over the years, I have tried many different solutions. Your chosen operating location will influence your choices. Operating from a picnic table in a park will give you more options. Remembering that limited lunchtime is a factor, you may want to just heave an end-fed wire up into a tree. An end-fed such as the commercial PAR ENDFEDZ or a home-made end-fed using a 31-foot length of wire to a 9:1 UNUN is perfect for this type of operation. I have also put up an end-fed where no trees were available by using a 31-foot Jackite pole as a temporary mast.

If your operations are going to be confined to a parking lot, your options are probably going to be more limited, but in no way does this mean you cannot be successful. I used Hamsticks on a mag-mount for years and was somewhat successful.

Just this past year, I was turned onto an even better solution by my good friend Bob Reisenweber, W3BBO. We both own Buddisticks. I had always used mine in the conventional manner, attaching it to a tripod or painter's pole, unwinding the counterpoise wire and going from there. Bob suggested a different approach. He was living in an apartment at the time and was limited to working from his car. He took

Photo C. Can't get away to a park to operate? Your company's parking lot might be an effective spot from which to operate, especially if you have a roof-mounted antenna.

Photo D. A Buddistick attached to a heavy-duty magnet mount can be a very effective antenna that's quick to set up and tear down. Also, no counterpoise is needed as long as your vehicle has a metal roof.

his Buddistick to his car in the lot and attached it to his car's mobile antenna mount (photo D). I tried it and had great success! Not only is it a quick and easy setup and tear-down, but by its very nature, the Buddistick is oriented towards multiband operation. Changing a tap on a coil when changing bands is so much easier than changing individual resonators, and the car provides a very nice ground plane for the antenna. No counterpoise wire is needed with this set up.

Selecting a Rig

As far as rig choice goes, there are many available to you as a QRPer. Just about any of the commercially offered QRP rigs, or any of those offered as kits, can be operated from the field quite easily. Toss one in a case, along with a key or microphone and a battery, and you are good to go. My weapon of choice is the Elecraft KX3. I use a Whiterook single-lever paddle, some earbuds, and a small 9800 mAH lithium-ion battery. It makes for a perfect portable station, light and easy to carry in a Lowepro 150 travel case (photo E).

The results have been fantastic! I can get into my car, break out my KX3, set up the Buddistick, and be on the air within mere minutes. For 40 meters, I don't tap the Buddistick's coil. For 20, 17, 15, 12, and 10 meters, I have found one tap position that allows the KX3's internal tuner to work effortlessly with the Buddistick.

Since last April, I have over 160 QSOs with this setup, making CW contacts using 5 watts or less with 48 different DX entities, 19 states, and 5 SOTA (Summits on the Air) peaks.

Photo E. The author's lunchtime-portable QRP station, built around his Elecraft KX-3 transceiver.

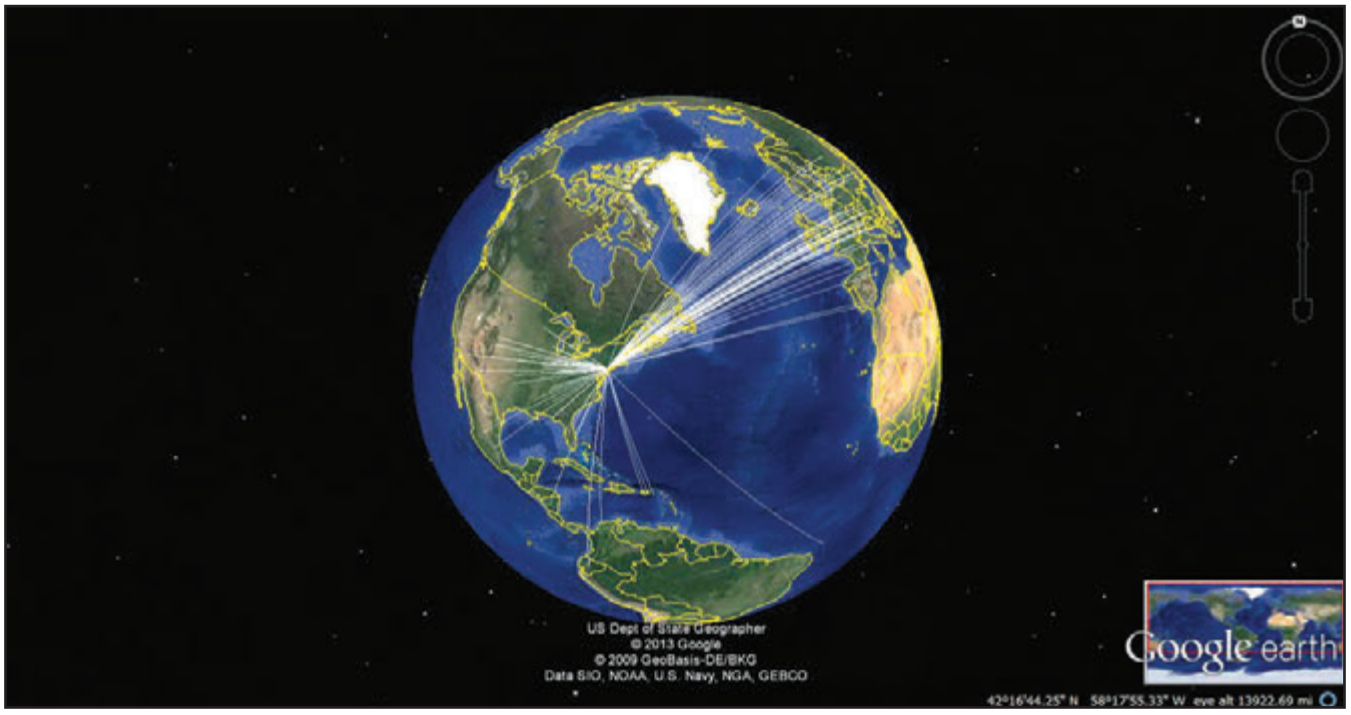

Photo F. A Ham Radio Deluxe overlay of W2LJ's 2013 lunchtime contact paths on a Google Earth® map. This view shows the contacts from a western hemisphere perspective.

Contacts are routinely made with stations 6,000 and 7,000 miles away, easily achieving the 1,000-mile-per-watt category. My most distant contact so far has been with R6YY at the eastern edge of European Russia at 8,522 miles. Photos F and G are screen-shots from my computer showing contact paths overlaid on Google Earth® maps by Ham Radio Deluxe.

Staying Sane

Even though lunchtime is limited, you can make a lot of contacts and more importantly you can have a lot of fun. I sometimes kid with my ham friends that if it weren't for lunchtime, I'd never get on the air at all. This hour has been my salvation. It supplies my daily needed fix of QRP, it breaks up my day, and it helps keep me sane. If you are in the same situation that I am, you should give this a try. You'll be glad you did and you just might be amazed by your results.

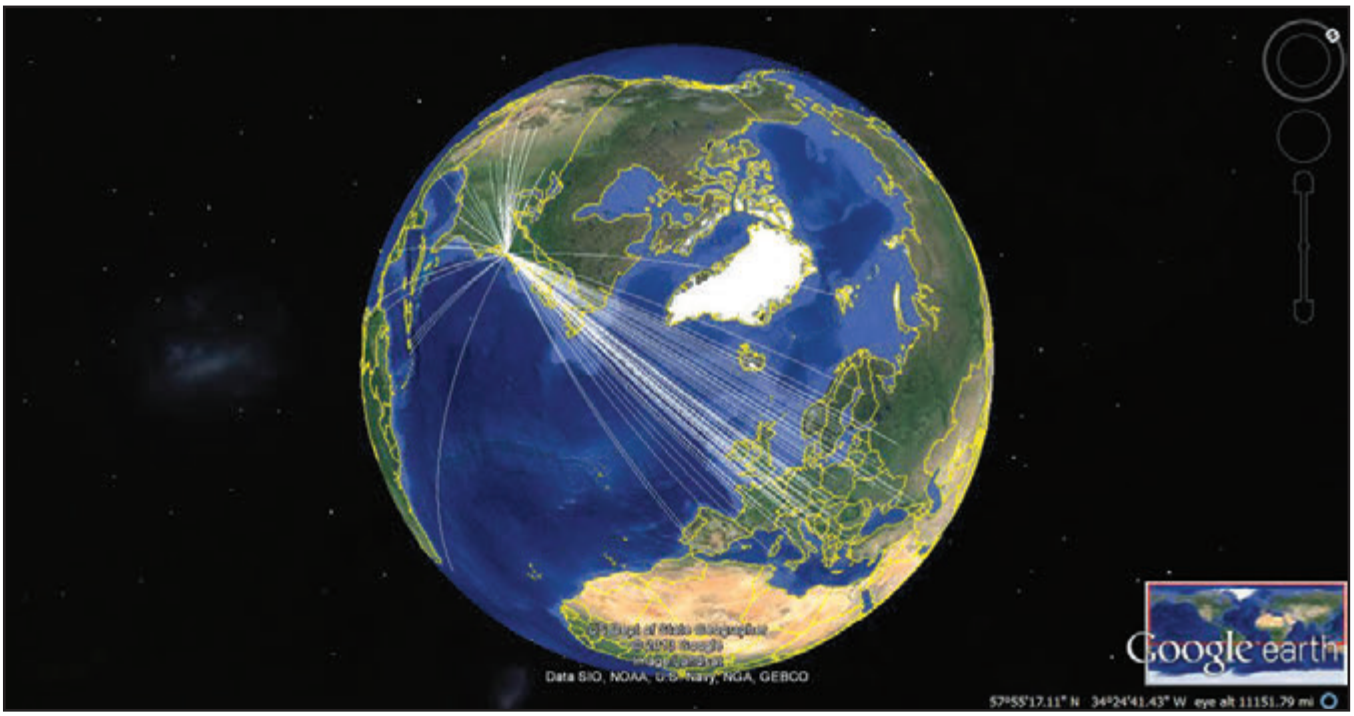

Photo G. The same contact paths as shown in photo F seen from an eastern hemisphere perspective. Five watts can travel quite a long way!

What at first seemed to the author like "a painful process" rutned into an adventure in learning while collecting contacts with all 50 states using just one-tenth of a watt.

Worked All States with 100 Milliwatts

BY JOHN WATKINS,* NØEVH

Working all 50 states with an accumulated power of 5 watts or an average of 1/10 watt for each state... I had heard of amateurs achieving this goal, but it sounded like a painful process that certainly one would not consider as part of an enjoyable hobby. But here I am…

It started innocently for me back in May of 2003 when I worked a killer-signal station in Michigan from my home state of Missouri on 40 meters. K8CV was so very strong into my QTH, and since Walt normally operated QRP, I thought the band must be in great shape. I quickly dropped my power to 1/10 of a watt and gave him a call. We continued our QSO with no problem.

With my courage up, I looked around 20 meters for a strong signal and worked a Virginia station within the hour, still at 100 milliwatts. Going to higher frequencies always helps the low-power enthusiast because of a better signal-to-noise ratio. I'm not sure whether I was lucky or if the bands were just amazing, but I began to watch for unusually strong signals from stations and to make note of how much power they were running.

Photo A. The author at his station in Missouri. His Elecraft KX-3 (lower right, behind pad) was his main "weapon" in contacting all 50 states at a power level of 1/10 of a watt.

Antenna Experiments

I recently had moved my vertical antenna from a ground mount to a chimney mount and noticed some improvement in performance. I did not have anything to compare it with, so I decided to install a horizontal 80-meter loop so that I had a choice between a vertical and a horizontal antenna. The average height of my loop was 25 feet, all I could muster in my yard. I began to compare antennas and was pleased to find that the loop was a killer antenna in close on 80 and 40, easily beating the vertical. Then on the upper bands—20 and above—the loop actually performed slightly better than the vertical. I about wore out my A/B antenna button the next few years doing comparisons. I feed the loop with 450-ohm ladder line back to the shack and through a 4:1 balun to my rig. As the experiments continued, my states total continued to grow.

I did not pursue the goal hard, just let it happen over the years whenever I heard great band conditions. Along the way, I hit a couple of fantastic openings and finally worked Hawaii and Alaska on 10 meters at the end of a contest when stations were hungry for contacts and had no answers to their CQs. You probably know the callsigns if you're a contester; those guys have excellent stations and copying skills. The 80-meter loop was my go-to antenna on about 90 percent of my Qs. So now I had the hard ones in the log, right?

Not really. The hard ones sometimes are the close ones or those with fewer ham radio stations on the air. Most of these states are close to my home QTH in Missouri (photo A), so I had a double-whammy. What to do?

I wanted to add 160 meters to my bands, so I went with the only antenna that would fit in my lot. It was an inverted L at 30 feet and the horizontal portion even had a dogleg in it because of my small lot. Thinking that I now had added a band, I was amazed when the inverted L began to beat out my 80-meter loop on some stations. I was feeding the L with a 9:1 balun, had two 80-foot above-ground radials attached at the ground feed point, and they ran around the house on the wooden hardscape. It was not ideal by the antenna book, but it sometimes beat the loop. I have now had the loop for ten years and the L for four. They split the very low-

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e-mail: <n0evh@sbcglobal.net>

power contacts at 50% each. I think it depends upon incoming/outgoing take-off angle of the signal.

The Hunt Begins in Earnest

Now that I had some success, I wanted to pursue some competitive stations and add to my state count. I felt the close-in states would be worked on 40 and 80 meters, so I needed to chase states during the winter months when thunderstorms were not raising the noise floor. Avoiding the QRN would go a long way to my signal being heard. Between my two best antennas—the loop and the inverted L—I figured that I could now present a decent signal on 80 through 10 meters. A few years passed and my total state count reached 40.

With the ARRL Sweepstakes contest coming in the fall of 2013, I checked the family calendar and discovered that I would have time to get on the air. Of my remaining ten states many were the small- or low-population ones. Checking the ARRL contest record website, I identified the top-scoring stations for those states and made my own personal hit list. I then set up CW Skimmer and married it to my Elecraft KX3 I/Q output so I could watch 48 kHz of whatever band I was on. This would give me an extra set of eyes trying to spot my targets. CW Skimmer produces a list that can be sorted by time, frequency, or callsign (see fig. 1), so when I took a stretch break, it was still watching for my targets. My final preparation step involved analyzing my logbook to see when I had heard my target states, on what bands and at what time of the day during the month of November. I now had a plan for working the Sweeps.

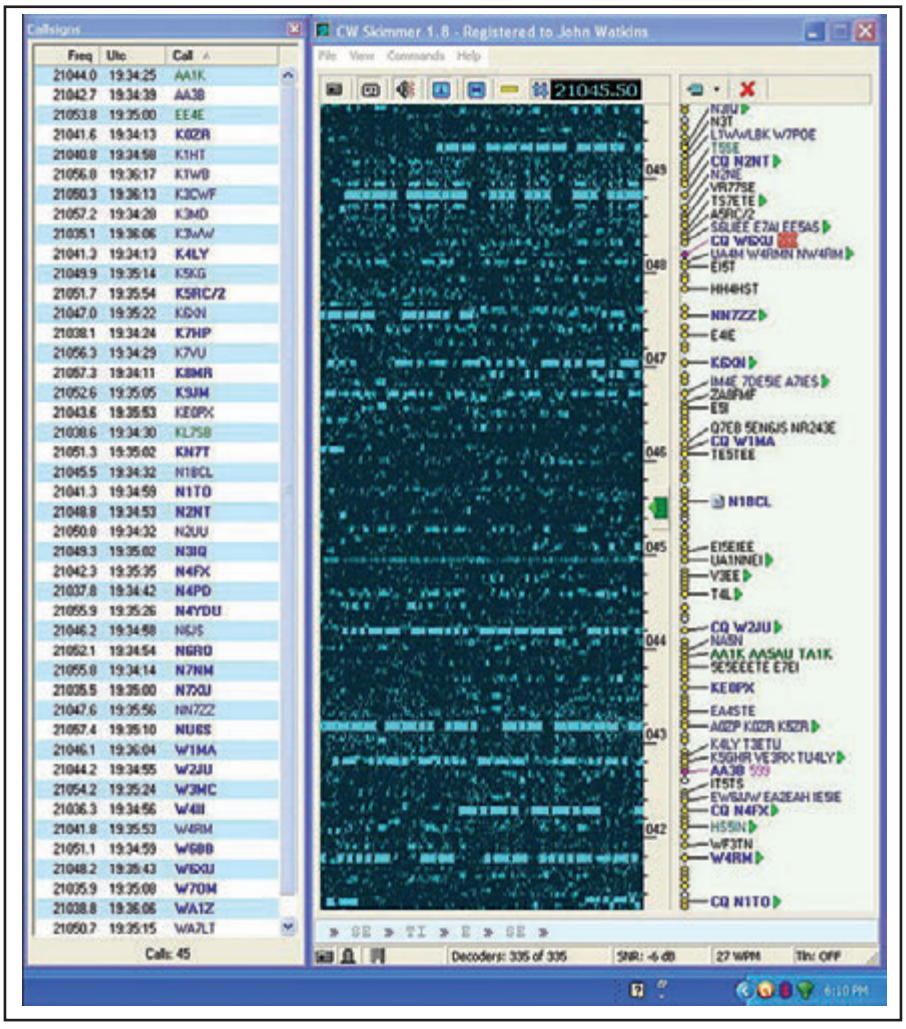


Fig. 1. Screen capture of CW Skimmer running on the right half of the screen. Taken during the heat of Sweepstakes action, you can see the list of stations that have been identified by the skimmer. They are sorted alphabetically by callsign.

Going All Out

On the first day of the contest, I jumped in with my 1/10 of a watt. In five hours I had logged eight of my ten target states with only two of them asking for repeats. Bands were in great condition and things were going very well. The fall weather and absence of thunderstorms across the nation were providing great signals and low QRN on 40 and 20 meters. States in the range of 400 to 600 miles—such as Indiana, Minnesota, and Oklahoma—were collected on 40. Kansas was logged on 80 in the early evening, before that band went long. The more distant states of New Mexico, Wyoming, and Montana were logged on 20.

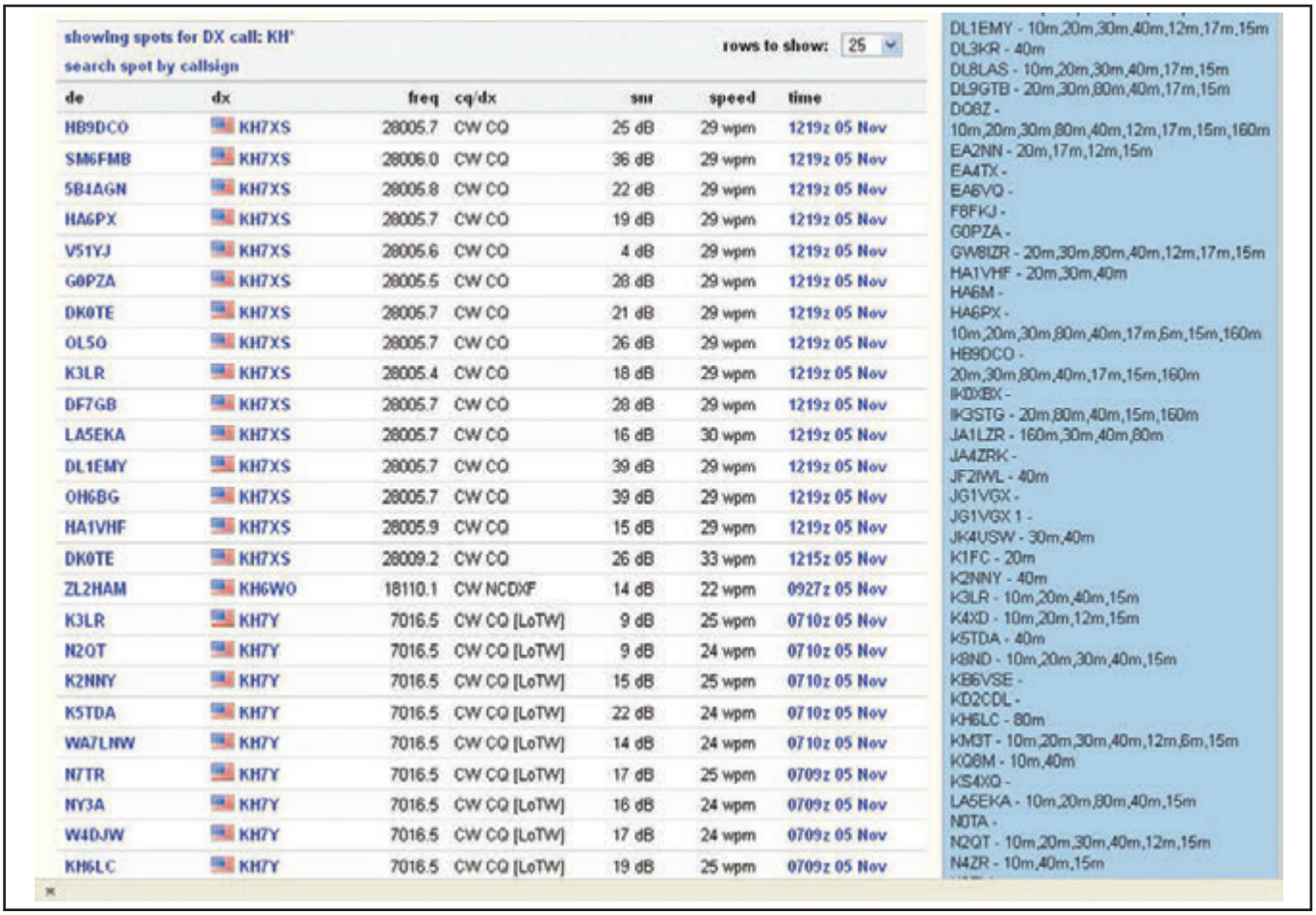
To be heard at these power levels you need to call in the clear (no one else calling the CQing station) and have great propagation and low noise at the other end. Now only West Virginia and Delaware were needed in my log, but reaching the East Coast from the Midwest was not going to work for me for the rest of the evening. While 40 meters was in great shape and signals to me were strong, the band was wall-to-wall signals, so the QRM was too intense for my little signal. I would have to wait for West Virginia and Delaware. There would be less signal congestion on Sunday morning, I reasoned, so I pulled the plug for the night.

The Sweeps event has 83 sections and many stations want to collect them all for a "clean sweep." Others are going for the sweep plus high score. Thus, the odds were in my favor that on the second day the sweep hunters would have moved on to other targets.

One great feature of CW Skimmer beyond the obvious is that it calculates the signal-to-noise ratio of the station tuned. I was running power 30 dB below most stations in the contest, so I was looking for signals that were at least 40 dB above my noise. Of course, too, I was hoping that the noise level was also low at their end. No point in calling a station that can't hear you. Once you have found your target station, listen to him/her work through the callers. Watch the ebb and flow of the action. Every few minutes the CQing station will get a brief lull in activity. No callers. That is your opportunity, so be ready with your memory keyer set at his speed and make it happen. Don't use your key, as this is not the time to make mistakes.

The Final Two

The next morning, West Virginia was in my log on 20 meters by 10 AM. I tried to get his attention for one-and-a-half hours. His signal was holding just fine at 45 dB over the noise, so once the competition had been worked and he had few callers, he was mine. A final snack from the kitchen and then back



showing spots for DX call: KH*
search spot by callsign

rows to show: 25

de	dx	freq	cq/dx	snr	speed	time
HB9DCO	KH7XS	28005.7	CW CQ	25 dB	29 wpm	1219z 05 Nov
SM6FMB	KH7XS	28006.0	CW CQ	36 dB	29 wpm	1219z 05 Nov
5B4AGN	KH7XS	28005.8	CW CQ	22 dB	29 wpm	1219z 05 Nov
HA6PX	KH7XS	28005.7	CW CQ	19 dB	29 wpm	1219z 05 Nov
V51YJ	KH7XS	28005.6	CW CQ	4 dB	29 wpm	1219z 05 Nov
G0PZA	KH7XS	28005.5	CW CQ	28 dB	29 wpm	1219z 05 Nov
DK0TE	KH7XS	28005.7	CW CQ	21 dB	29 wpm	1219z 05 Nov
OL5O	KH7XS	28005.7	CW CQ	26 dB	29 wpm	1219z 05 Nov
K3LR	KH7XS	28005.4	CW CQ	18 dB	29 wpm	1219z 05 Nov
DF7GB	KH7XS	28005.7	CW CQ	28 dB	29 wpm	1219z 05 Nov
LA5EKA	KH7XS	28005.7	CW CQ	16 dB	30 wpm	1219z 05 Nov
DL1EMY	KH7XS	28005.7	CW CQ	39 dB	29 wpm	1219z 05 Nov
OH6BG	KH7XS	28005.7	CW CQ	39 dB	29 wpm	1219z 05 Nov
HA1VHF	KH7XS	28005.9	CW CQ	15 dB	29 wpm	1219z 05 Nov
DK0TE	KH7XS	28009.2	CW CQ	26 dB	33 wpm	1215z 05 Nov
ZL2HAM	KH6WO	18110.1	CW NCDXF	14 dB	22 wpm	0927z 05 Nov
K3LR	KH7Y	7016.5	CW CQ [LoTW]	9 dB	25 wpm	0710z 05 Nov
N2QT	KH7Y	7016.5	CW CQ [LoTW]	9 dB	24 wpm	0710z 05 Nov
K2NNY	KH7Y	7016.5	CW CQ [LoTW]	15 dB	25 wpm	0710z 05 Nov
K5TDA	KH7Y	7016.5	CW CQ [LoTW]	22 dB	24 wpm	0710z 05 Nov
WA7LNW	KH7Y	7016.5	CW CQ [LoTW]	14 dB	24 wpm	0710z 05 Nov
N7TR	KH7Y	7016.5	CW CQ [LoTW]	17 dB	25 wpm	0709z 05 Nov
NY3A	KH7Y	7016.5	CW CQ [LoTW]	16 dB	24 wpm	0709z 05 Nov
W4DJW	KH7Y	7016.5	CW CQ [LoTW]	17 dB	24 wpm	0709z 05 Nov
KH6LC	KH7Y	7016.5	CW CQ [LoTW]	19 dB	25 wpm	0709z 05 Nov

DL1EMY - 10m,20m,30m,40m,12m,17m,15m
DL3KR - 40m
DL8LAS - 10m,20m,30m,40m,17m,15m
DL9GTB - 20m,30m,80m,40m,17m,15m
DO8Z - 10m,20m,30m,80m,40m,12m,17m,15m,160m
EA2NN - 20m,17m,12m,15m
EA4TX -
EA6VQ -
F8FKJ -
G0PZA -
GW8IZR - 20m,30m,80m,40m,12m,17m,15m
HA1VHF - 20m,30m,40m
HA6M -
HA6PX - 10m,20m,30m,80m,40m,17m,6m,15m,160m
HB9DCO - 20m,30m,80m,40m,17m,15m,160m
IK0XBX -
IK3STG - 20m,80m,40m,15m,160m
JA1LZR - 160m,30m,40m,80m
JA4ZRK -
JF2IWL - 40m
JG1VGX -
JG1VGX 1 -
JK4USW - 30m,40m
K1FC - 20m
K2NNY - 40m
K3LR - 10m,20m,40m,15m
K4XD - 10m,20m,12m,15m
K5TDA - 40m
K8ND - 10m,20m,30m,40m,15m
KB6VSE -
KD2CDL -
KH6LC - 80m
KM3T - 10m,20m,30m,40m,12m,6m,15m
KQ8M - 10m,40m
KS4XQ -
LA5EKA - 10m,20m,80m,40m,15m
N0TA -
N2QT - 10m,20m,30m,40m,12m,15m
N4ZR - 10m,40m,15m

Fig. 2. Screen capture from RBN (Reverse Beacon Network) website showing a search on KH (* is a wildcard) for this search engine. This provides one method for finding active stations from a given part of the world or targeting a specific station with its callsign.*

to the bands. The only Delaware station I had spotted was doing search-and-pounce, so my progress was stalled. I scouted 10 meters for about an hour, but it became obvious that the east and west coast stations were working each other, not the "fly-over" states. I then dropped down to 15 meters and found that my Delaware target station had begun to call CQ. Got him in just a couple of minutes!

Ten years, three radios, and multiple antenna experiments, what fun it was along the way! And that tough Delaware station I finally found and logged; the morning after the Sweeps I got my confirmation for the QSO via Logbook of the World (LoTW)! Got to love electronic logging! For those who like stats, here are states per band: 10m-6, 12m-0, 15m-6, 17m-1, 20m-20, 30m-1, 40m-14, 80m-2, 160-0.

There were some interesting highlights recorded in my log during my 10-year adventure. Three states were worked when I was calling CQ with 1/10 of a watt. While working in the shack on a project I would put the rig on a slow CQ repeat and listen on an open frequency. Five additional states were logged while they were running 1000 watts, 40 dB above my station. They heard me on my first call in the clear. When Mother Nature gives you the reflection, you just have to be there! About half of my states were worked during casual non-contest periods, and it was always fun to chat with the other ham about low-power radio. One old timer said he had never worked a station at a tenth of a watt and asked me to stand by while he turned his QRO rig down to 5 watts to give QRP a try. Another ham in Washington had been a World War II Army radio operator and carried radio gear on mules. You just never know whom you will run into on the bands.

Your odds of being on the right band for those great reflections can be increased by using the *Reverse Beacon Network* (fig. 2). If you are collecting states and are not restricted by assisted rules in a contest, you simply do a search on KH* if looking for a Hawaiian station or KL* if Alaska is not in your log. Of course, in my case I was looking for known big-station contesters, so I searched their specific callsigns. No need to fish where there are no fish.

Preparing that target station list was also valuable because several big contest stations had callsigns out of their call districts. For example, I would have never taken a second look at an 8 station while searching for my 50th state of Delaware.

What's Next?

So what to do now? Well, I am still working on DXCC at 1 watt. Guess I will close that out this winter, then maybe start on DXCC at 1/10 of a watt, but oh, I already have. Along the way while working on my states goal, I worked 17 DX entities at a 1/10 of a watt, so here we go again!

By the way, as an extra benefit of running very low power, you will end up with some great antennas! The ones that are marginal will have been taken down and redesigned by the time you get your 50 states in the logbook. Give it a go and enjoy the journey at whatever power level you choose to set as your goal.

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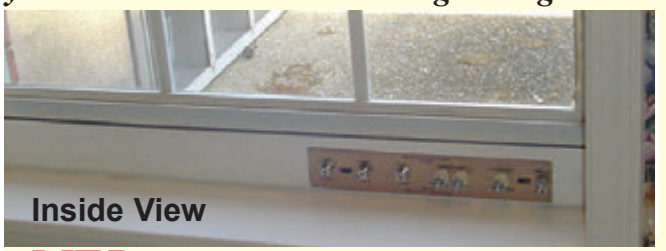
Weather-proof window feedthrough panels bring coax, balanced lines, HF/VHF/UHF antennas, random wire antennas, ground, rotator/antenna switch cables and DC/AC power into your hamshack without drilling through walls!



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Simply place in window sill and close window. One cut customizes it for any window up to 48 inches. Use horizontally or vertically. Connectors are mounted on inside/outside stainless steel plates and attached to a 4 foot long, $3^{1}/_{2}$ inch high, $^{3}/_{4}$ inch thick *pressure-treated* wood panel. Has excellent insulating properties. Weather-sealed with a heavy coat of long-lasting white outdoor enamel paint. Edges sealed by weather-stripping. Seals and insulates against all weather conditions. Includes window locking rod.

Inside/outside stainless steel plates ground all coax shields. Stainless steel ground post brings ground in.

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Four 50 Ohm *Teflon*(R) SO-239 coax connectors lets you feed HF/VHF/UHF antennas at full legal power limit.

A 50 Ohm *Teflon*(R) coax *N-connector* lets you use any antenna up to 11 GHz, including 450 MHz, UHF, satellite, moon bounce and 2.4/5.8 GHz Wi-Fi antennas.

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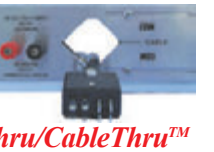
MFJ's exclusive *Adaptive Cable Feedthru*™ lets you bring in rotator/antenna switch cable, etc. without removing connectors (up to $1^{1}/_{4}$X$1^{5}/_{8}$ in). Adapts to virtually *any* cable size. Seals out rain, snow, adverse weather.



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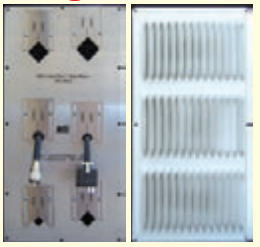
Bring cables thru eave of your house



MFJ-4616 shown with standard full-size vent (not included) it replaces. For *6 Cables* **$26.95**

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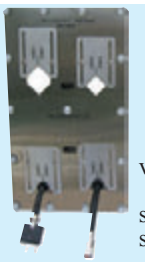
Bring in coax, rotator, antenna switch, power cables, etc. *with connectors* up to $1^{1}/4$x$1^{5}/8$ inches!

Sliding plates and rubber grommets adjust for virtually any cable size to seal out adverse weather, insects and varmints. Use existing vent hole, mounting screws and screw holes.

AdaptiveCable™ Wall Plates



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Bring nearly any cable -- rotator, antenna switch, coax, DC/ AC power, etc. -- through walls *without removing connectors* (up to $1^{1}/_{4}$x$1^{5}/_{8}$ inches). Sliding plates and rubber grommets adjust hole size to weather-seal virtually any size cable.

Includes *stainless steel* plates for each side of wall, sliding plates, rubber grommets, weather stripping and screws.

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Announcing:

The 2014 CQ World-Wide WPX Contest

SSB: March 29–30, 2014 CW: May 24-25, 2014
Starts: 0000 UTC Saturday Ends: 2359 UTC Sunday

I. OBJECTIVE: For amateurs world wide to contact as many amateurs and prefixes as possible during the period of operation.

II. PERIOD OF OPERATION: 48 hours. Single Operator stations may operate 36 of the 48 hours – **off times must be a minimum of 60 minutes** during which no QSO is logged. Multi-operator stations may operate the full 48 hours.

III. BANDS: Only the 1.8, 3.5, 7, 14, 21, and 28 MHz bands may be used. **Observance of established band plans is strongly encouraged.**

IV. EXCHANGE: RS(T) report plus a progressive contact serial number starting with 001 for the first contact. Note: Multi-Two and Multi-Unlimited entrants use separate serial number sequences on each band.

V. SCORING:

A. Score: The final score is the result of the total QSO points multiplied by the number of different prefixes worked.

B. QSO Points: A station may be worked once on each band for QSO point credit:

1. Contacts between stations on different continents are worth three (3) points on 28, 21, and 14 MHz and six (6) points on 7, 3.5, and 1.8 MHz.

2. Contacts between stations on the same continent, but different countries, are worth one (1) point on 28, 21, and 14 MHz and two (2) points on 7, 3.5, and 1.8 MHz. Exception: For North American stations only—contacts between stations within the North American boundaries (both stations must be located in North America) are worth two (2) points on 28, 21, and 14 MHz and four (4) points on 7, 3.5, and 1.8 MHz.

3. Contacts between stations in the same country are worth 1 point regardless of band.

C. Prefix Multipliers: The prefix multiplier is the number of valid prefixes worked. Each PREFIX is counted only once regardless of the band or number of times the same prefix is worked.

1. A PREFIX is the letter/numeral combination which forms the first part of the amateur call. Examples: N8, W8, WD8, HG1, HG19, KC2, OE2, OE25, LY1000, etc. Any difference in the numbering, lettering, or order of same shall count as a separate prefix. A station operating from a DXCC entity different from that indicated by its call sign is required to sign portable. The portable prefix must be an authorized prefix of the country/call area of operation. In cases of portable operation, the portable designator will then become the prefix. Example: N8BJQ operating from Wake Island would sign N8BJQ/KH9 or N8BJQ/NH9. KH6XXX operating from Ohio must use an authorized prefix for the U.S. 8th district (/W8, /AD8, etc.). Portable designators without numbers will be assigned a zero (Ø) after the second letter of the portable designator to form the prefix. Example: PA/N8BJQ would become PAØ. All calls without numbers will be assigned a zero (Ø) after the first two letters to form the prefix. Example: XEFTJW would count as XEØ. Maritime mobile, mobile, /A, /E, /J, /P, or other license class identifiers do not count as prefixes.

2. Special event, commemorative, and other unique prefix stations are encouraged to participate. Prefixes must be assigned by the licensing authority of the country of operation.

VI. ENTRY CATEGORIES:

A. Single Operator Categories: One person (the operator) performs all operating and logging functions. There is no limit on band changes. Only one transmitted signal is permitted at any time.

1. Single Operator: QSO alerting assistance of any kind is prohibited (see rule IX.2).

a. High Power (All Band or Single Band): Total output power must not exceed **1500 watts.**

b. Low Power (All Band or Single Band): Total output power must not exceed **100 watts.**

c. QRP (All Band or Single Band): Total output power must not exceed **5 watts.**

2. Single Operator Assisted: Entrants in this category may use QSO alerting assistance (see rule IX.E).

a. High Power Assisted (All Band or Single Band): Total output power must not exceed **1500 watts.**

b. Low Power Assisted (All Band or Single Band): Total output power must not exceed **100 watts.**

c. QRP Assisted (All Band or Single Band): Total output power must not exceed **5 watts.**

B. Single Operator Overlay Categories: Single Operator entrants may **also** submit their log for **one** of the categories shown below by adding an additional line in the Cabrillo log file header called CATEGORY-OVERLAY. All Overlay entries are grouped into high power and low power in the results.

1. Tribander/Single Element (TB-WIRES): During the contest an entrant shall use only one (1) tribander (any type, with a single feed line from the transmitter to the antenna) for 10, 15, and 20 meters and single-element antennas on 40, 80, and 160 meters.

2. Rookie (ROOKIE): To enter this category the operator must have been licensed as a radio amateur three (3) years or less on the date of the contest The date first licensed **must** be indicated in the SOAPBOX field. An operator who has won a plaque in any previously published CQ contest is not eligible for the Rookie category.

C. Multi-Operator Categories (All Band only): More than one person can contribute to the final score during the official contest period. Select category based on number of transmitted signals.

1. Single-Transmitter (MULTI-ONE): Only one transmitted signal is permitted at any time. A maximum of ten (10) band changes may be made in any clock hour (00 through 59 minutes). For example, a change from 20 meters to 40 meters and then back to 20 meters counts as two band changes. Use a single serial number sequence for the entire log.

a. High Power. Total output power of each transmitted signal must not exceed **1500 watts.**

b. Low Power. Total output power of each transmitted signal must not exceed **100 watts.**

2. Two-Transmitter (MULTI-TWO): A maximum of two transmitted signals is permitted at any time on two different bands. Both transmitters may work any station. A station may only be worked once per band regardless of which transmitter is used. **The log must indicate which transmitter made each QSO** (column 81 of CABRILLO QSO template for CQ contests). Each transmitter may make a maximum of eight (8) band changes in any clock hour (00 through 59 minutes). Use a separate serial number sequence for each band. Total output power of each transmitted signal must not exceed **1500 watts.**

3. Multi-Transmitter (MULTI-UNLIMITED): A maximum of six transmitted signals, one per band, at any one time. Six bands may be activated simultaneously. Use a separate serial number sequence for each band. Total output power of each transmitted signal must not exceed **1500 watts.**

D. Checklog: Entry submitted to assist with the log checking. The entry will not have a score in the results and the log will not be made public.

VII. AWARDS: A single-band log will be eligible for a single-band award only. To be eligible for an award, a Single Operator station must show a minimum of 4 hours of operation. Multi-operator stations must operate a minimum of 8 hours.

A. Plaques are awarded to recognize top performance in a number of categories. View the current list of plaques and sponsors at <www.cqwpx.com/plaques.htm>.

Only one plaque will be awarded per entry. A station winning a World plaque will not be considered for a sub-area award. That award will be given to the runner-up for that area if the number of entries justifies the award.

B. Certificates will be awarded to the highest scoring station in each category listed under Section VI:

1. In every participating country.

2. In each call area of the United States, Canada, Russia and Japan.

3. At the discretion of the contest director second- and third-place awards may be made.

VIII. CLUB COMPETITION: A plaque will be awarded each year to the club that has the highest aggregate score from logs submitted by its members. To be listed in the results, a minimum of three logs must be received from a club.

A. The club must be a local group and not a national organization (e.g., ARRL or DARC).

B. Participation is limited to members residing in or operating from a local geographic area (except for DXpeditions conducted by members living within the defined club geographic area). Club contributions from DXpedition scores are a percentage of the number of club members on the DXpedition.

C. Single-operator entries can only contribute to one club. Multi-operator scores may be allocated to multiple clubs as a percentage of the number of club members participating in the operation. A minimum of four logs must be received for a club to be listed in the results. Checklog entries are not counted for the club score.

IX. DEFINITIONS OF TERMS:

1. Station location: The area in which all the transmitters, receivers and antennas are located. All transmitters and receivers must be within a single 500-meter diameter circle. Antennas must be physically connected by RF transmission lines to the transmitters and receivers.

2. QSO alerting assistance: The use of any technology or other source that provides call sign or multiplier identification along with frequency information to the operator. It includes, but is not limited to, use of DX cluster, packet, local or remote call sign and frequency decoding technology (e.g., CW Skimmer or Reverse Beacon Network), or operating arrangements involving other individuals.

X. GENERAL RULES FOR ALL ENTRANTS:

A. Entrants must operate within the limits of their chosen category when performing any activity that could affect their submitted score.

B. A different callsign must be used for each entry. Only the entrant's callsign may be used to aid the entrant's score.

C. Do not exceed the total output power limitation of the chosen category on any band. Total output power on any band at any time is measured at the final output of the active amplifier(s).

D. Self-spotting or asking to be spotted is not permitted.

E. All operation must take place from one station location. Remote operation is permitted if the physical location of all transmitters, receivers, and antennas are at one station location. A remotely operated station must obey all station license, operator license, and category limitations.

F. Remote receivers outside the station location are not permitted.

G. Only one signal on a band is allowed at any time. When two or more transmitters are present on the same band, a hardware device MUST be used to prevent more than one signal at any one time. Alternating CQs on two or more frequencies on a band is not permitted.

H. All requests for contacts, responses to calls, and copying of call signs and contest exchanges must be accomplished during the contest period using the mode and frequencies of the contest.

I. Correction of logged call signs and exchanges after the contest by using any database, recordings, email or other methods of confirming QSOs is not allowed.

X. LOG INSTRUCTIONS: Electronic submission of logs is **required** for all entrants who use a computer to log the contest or prepare contest logs.

A. The log MUST show the following for each contact: correct date and time in UTC, frequency (or band), call sign of the station worked, exchange sent, and exchange received. A log without all required information may be reclassified to Checklog. Contacts should be logged immediately as they are completed. Stations competing for World and Continent awards should provide accurate frequencies for all contacts in the log.

B. Single band entrants are required to include all contacts made during the contest period, even if on other bands. Only contacts made on the band specified in the Cabrillo header or summary sheet will be considered for scoring purposes. Logs with contacts only on one band will be classified as single band entries.

C. The CABRILLO file format is the standard for logs. See <www.cqwpx.com/lcabrillo.htm> for detailed instructions on filling out the CABRILLO file header. Failure to fill out the header correctly may result in the entry being placed in the wrong category or reclassified as a Checklog. Note: U.S. stations must indicate the station location in the CABRILLO header (e.g., LOCATION: OH).

D. E-mail or Web upload is the expected method of log submission. SSB logs in CABRILLO format should be sent to <ssb@cqwpx.com>. CW logs in CABRILLO format should be sent to <cw@cqwpx.com>. Include only the entry call sign in the "Subject:" line of the e-mail. Web upload of logs is available at <www.cqwpx.com/logcheck/>. All logs received will be confirmed via e-mail. A listing of logs received can be found at <www.cqwpx.com>.

E. Instructions for NON-CABRILLO electronic logs: If you are not able to submit a CABRILLO format log, please contact the Contest Director for assistance with submitting another format.

F. Instructions for paper logs: Paper logs may be mailed to CQ WPX Contest, P.O. Box 481, New Carlisle, OH 45344, USA. Each paper log entry must be accompanied by a Summary Sheet listing all scoring information, the category of competition, and the entrant's name and mailing address in BLOCK LETTERS.

XI. LOG DEADLINE:

A. All entries must be emailed or postmaked WITHIN FIVE (5) DAYS after the end of the contest: SSB logs no later than 2359 UTC 4 April 2014, CW logs no later than 2359 UTC 30 May 2014 Resubmitting an entry after the deadline will result in it being considered as a late log.

B. An extension may be requested by e-mail to <director@cqwpx.com>. The request must state a legitimate reason and must be received before the log deadline. Extensions are granted only upon confirmation by the Contest Director.

C. Logs submitted or postmarked after the deadline may be listed in the results, but are not eligible for awards.

XII. JUDGING: The CQ WPX Contest Committee is responsible for checking and adjudicating the contest entries. Entrants are expected to follow the rules and best amateur radio practices. Violation of the rules of the contest or unsportsmanlike conduct may lead to disciplinary action by the Committee.

A. Unsportsmanlike Conduct: Examples of unsportsmanlike conduct include, but are not limited to:

1. ARRANGING or CONFIRMING any contacts during the contest by use of ANY non-amateur radio means such as telephones, Internet, instant messaging, chat rooms, VoIP, social media or web sites.

2. Transmissions by the entrant on frequencies outside of license limitations.

3. Changing times in the log to meet band change or off time rules.

4. Taking credit for excessive unverifiable QSOs or unverifiable multipliers.

5. Signals with excessive bandwidth (e.g., splatter, clicks) or harmonics on other bands.

6. Running stations making more than 3 contacts without sending their callsign.

B. Observer Program: The Committee may request of any entrant to accept a visit by an observer during the contest. Failure to allow a Committee appointed observer full access to the station during the contest period may result in the entry being disqualified.

C. Disciplinary Actions: In the event of a violation, the entrant is subject to disqualification at the discretion of the Committee.

1. DISQUALIFICATION: Entry will be listed at the end of the published results and is not eligible for an award.

2. Notification of Committee actions will be sent by email to the address provided with the log submission. The entrant has five days to appeal the decision to the Contest Director. After that time the decision is final.

3. An entrant may withdraw the submitted log for any reason within 30 days of the log deadline. Contact the Contest Director for instructions.

4. The Committee reserves the right to change the category of any entry based on its examination of the log or other information.

D. Log Checking: All logs are checked using custom software and human judgment.

1. Duplicate contacts are removed with no additional penalty.

2. Contacts with an incorrectly received exchange are removed with no additional penalty.

3. Call sign errors or call signs not found in the other log are removed and receive a penalty of two times the QSO point value for that contact.

4. Contacts that do not meet the band change rules for multi-operator entries are removed with no additional penalty.

XIII. DECLARATION: By submitting a CQ WPX Contest log, and in consideration of the efforts of the CQ WPX Contest Committee to review and evaluate that log, an entrant unconditionally and irrevocably agrees that he/she has: 1) read and understood the rules of the contest and agrees to be bound by them, 2) operated according to all rules and regulations that pertain to amateur radio for the station location, 3) agreed the log entry may be made open to the public, and 4) accepted that the issuing of disqualifications and other decisions of the Committee are official and final. If an entrant is unwilling or unable to agree to all of the foregoing, the entrant should not submit the entry or submit the entry as a Checklog only.

Questions pertaining to the CQ WPX Contest may be emailed to the CQ WPX Contest Director at <director@cqwpx.com>.

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You may have heard about JT-65 for working moonbounce and meteor scatter. But what about JT-9, for working QRP on HF? Here's an introduction…

JT9: QRP with the Magic Multiplied

BY GARY LILJEGREN,* W4GAL

Operating with 100 watts of output power and bouncing a signal off the ionosphere when talking with stations thousands of miles away still seems amazing to me. I've done that for a long time. However, when I reduce the power level to one one-hundredth of that, and have many QSOs world-wide (photo A), that's beyond amazing; it's magic.

There is a relatively new mode available today that offers an opportunity for *multiplied magic*. The mode is *JT9,* created by Joe Taylor, K1JT, and friends. JT9 has been running on the bands a little more than a year. It functions a bit like PSK with an audio signal applied to SSB (upper sideband) in the CW portion of the bands.

Five Reasons Why the Magic is Multiplied

1. Surprising success operating QRP with results as if the "ERP" is considerably greater
2. Talking to stations that are below the noise level; they are inaudible in the speaker
3. The ability to easily have QSOs from any language to any other language
4. A perfect mode for a highly successful operation as a temporary or portable station
5. Visualizing the international reception results of your signal on a world map

Let's explain the magic of the five reasons:

1) FM and TV broadcast stations talk about their *effective radiated power* (ERP) in watts. The signal has been "amplified" at the receiving locations by the use of gain antennas at the transmitter. The ERP comparison is similar to JT9, but for a different reason. A near-broadcast-quality AM signal in ham radio is about 6,000 Hz wide. All of the transmitting power is going into that 6-kHz signal. On SSB, the signal is somewhere around 2,800 Hz wide and thus the ERP for SSB is maybe twice that of an AM signal. (An aside to you technically astute hams reading this: cut me a little slack on the technical specifics.) Without comparing other modes for efficiency, let's look at JT9. The width of the signal is about 16 Hz. That means the signal bandwidth is less than 1/100th that of an SSB signal. One watt on JT9 does somewhere

*e-mail: <gary@liljegren.com>

Photo A. The author's impressive station appears to have nearly as many computers in it as radios. (Photos by the author)

Photo B. Shirl Regis, W9OMG, operating JT9 at 5 watts with great results from her home in North Florida using an ICOM IC-718 and a temporary vertical antenna.

around the same job communicating information as a 100-watt SSB signal. QRP, enhanced! And this gets us to the next magic point.

2) When you hear the description of how JT9 works with very-weak stations (below), it won't sound possible at first to copy a station that is below the noise level. What Joe has done is apply the technology of weak-signal moonbounce and meteor-scatter data acquisition on UHF and VHF to the HF and MF bands. JT9 is intended for the latter two. He has created a way of transmitting and receiving minute changes of audio frequency to form letters and symbols. The transmitted signal is a steady tone but the audio pitch varies in frequency at specified intervals, depending on what the letter is. The WSJT-X software detects those audio changes when copying the transmission. The software also computes the received signal strength of the other station. When a contact is made, the software is brutally honest and displays the signal strength on the display operating panel in dBs. Joe says that if the signal is –15 or less, nothing is heard through the speaker; the received signal is inaudible. However, the software can copy it! I regularly talk with stations whose signals are at the –20 or –25 level. It is as if the noise is moving around and the software simply copies the letters from it. Magic. Multiplied.

3) If English is not your first language, or not one of your languages at all, this is an ideal mode. It is better than PSK, RTTY, and certainly better than SSB. It is the same as a brief contest contact in CW, in which the exchange is language-independent. With JT9, the QSO components are already established. Any language can communicate with any other language because the software arranges the QSO. Translating the explanation of JT9 into any language can cover the setup and operation of the WSJT-X software and you are on the air with JT9. K1JT has given permission to translate the WSJT-X User's Guide into any language as long as he is sent a copy of the translation so he can add it to the website.

4) What if you are a college student with a shiny new license, a limited ham radio budget, and questionable antenna possibilities? Or what if you just love seeing what you can do with low power and temporary antennas? With JT9, you can call CQ with 1 or 2 or 5 watts and get responses! You can have great success even over thousands of miles. This mode is perfect when you have a

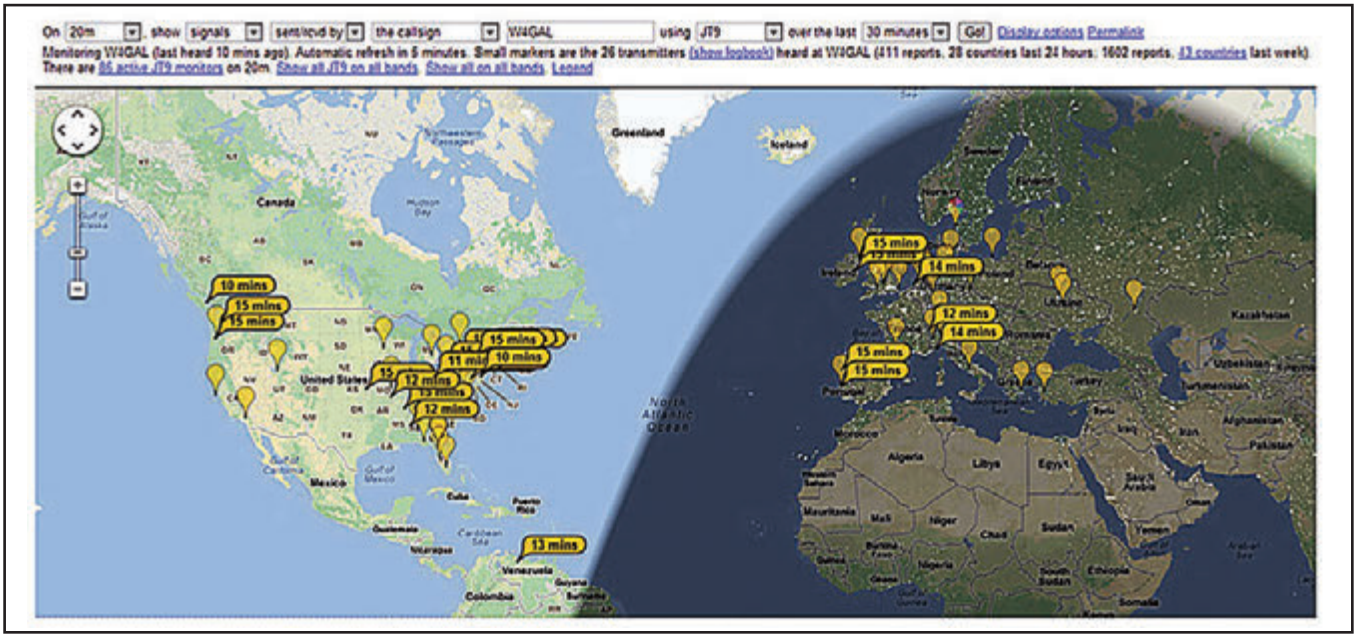

Fig. 1. Screen shot of stations hearing the author's 1-watt JT-9 signal from Florida, as displayed on the PSKREPORTER website.

little-pistol station. Photo B shows Shirl Regis, W9OMG, operating JT9 QRP with great results from her home in North Florida using an ICOM IC-718 cranked down to 5 watts and a temporary vertical antenna. Even if you have a big-gun station, see what you can do with 1 watt. Put the "wow" factor back into ham radio.

5) There is a powerful website available to us which displays signal reception reports from locations worldwide for a variety of modes. JT9 is one of the modes, and PSKREPORTER[1] is the site. After transmitting, you can look at PSKREPORTER and see which of the volunteer reporting stations received your transmission. I am having great success running somewhere between 1 and 5 watts, and I frequently look to see which stations, in which other continents, have also copied my signal.

Here is an example: I was using my TS-590 on a JT9 QSO with 1 watt power output through about 70 feet of RG8X to my vertical antenna in North Florida. I waited a few minutes after the QSO to do a "PrintScreen" because it takes a little while for the reporting stations to populate the PSKREPORTER map. Where was my transmission heard? Look at fig. 1. Holy smoke! My 1-watt

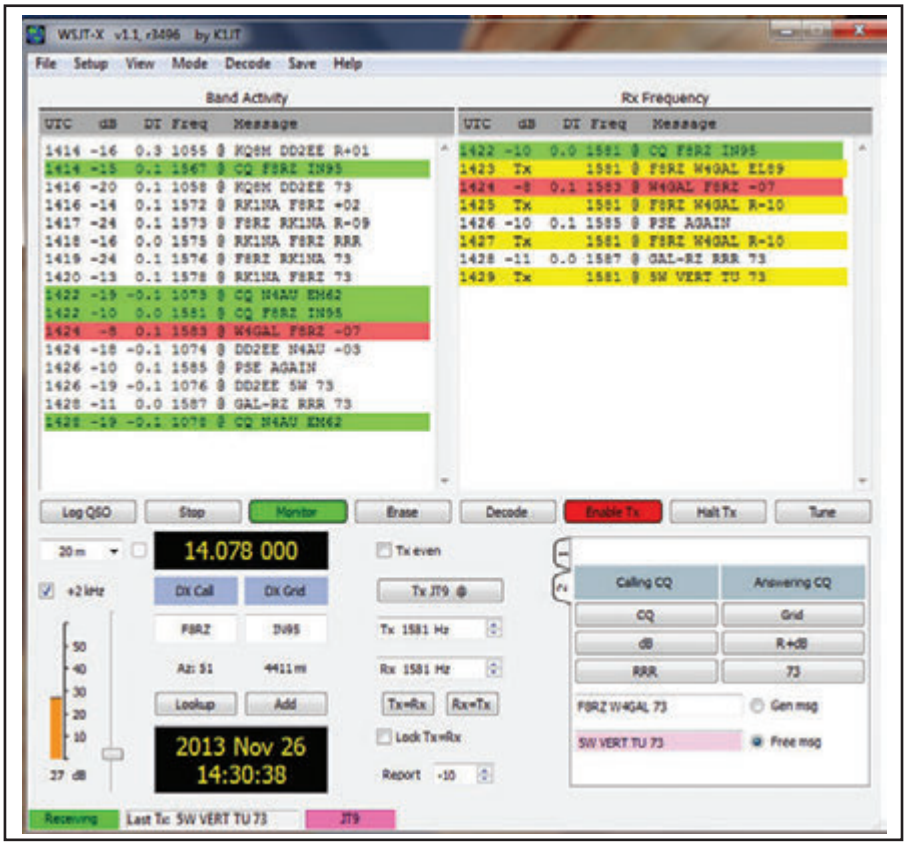

Fig 2. JT-9 software screen, showing band activity and a QSO between W4GAL and F8RZ. See text for details.

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signal was heard across North America and into South America and Europe at that time of day.

How a QSO Works

Fig. 2 illustrates a JT9 QRP QSO with Jean Raynaud, F8RZ, in France. To explain the lines on the Band Activity screen, Jean called CQ (CQs appear in green). He was answered by RK1NA in Russia. At the end of their QSO, Jean called CQ again and I answered by double-clicking on his CQ. That started our QSO in the Rx Frequency window. He replied to me (his reply appeared in red) and said I was a –07 in signal strength. I replied that he was a –10. The software supplies the data automatically. He asked for a repeat and I sent it again. He acknowledged it and said 73. I said I was running 5 watts with a vertical, and thank you 73. End of QSO.

Here's the rub: It takes 50 seconds to send each 13-character message—which, I might add, gives you enough time to enjoy your cup of coffee and to look at QRZ.com and see who you are talking with during the QSO.

Equipment

You need a transceiver running 20 watts or less power, a computer, a sound-card interface to connect the two, and some free software. The standard SignaLink and Rigblaster types of modems work fine, and many rigs like my Kenwood TS-590 have a built-in sound card which does the job. I like using an outboard unit instead because it has a knob on it so I can adjust the drive level in real-time and choose the output power as I go. I can set the TS-590 at its minimum 5-watt level and dial-down the drive to where the output showing on my MFJ antenna tuner is 1 watt.

You will also need to sync your computer clock with a timing program such as Meinberg NTP.[2] If you can't get your computer to within one-half second of an international time standard, the WSJT-X program may not decode anyone. You can find what you need to know with instructions and downloadable software on the WSJT-X website.[3] Read the User's Guide first; don't wait until you have problems! And you can get help from hams experienced with JT9 if needed.[4]

Success, PLUS

When operating from your apartment with 5 watts and a stealth antenna hanging out the window, you can have amazing success on JT9. And it's true pretty much worldwide. You can call CQ and you will get responses. On JT9, if you're running 1 watt and not getting answered by the station you are calling but you can see on PSKREPORTER that you were "heard" by 10 others, it's not discouraging at all. You will also get a lot of eQSLs if you enjoy that. There seems to be a direct correlation between using a computer in your ham operating mode and the percentage of contacts confirming the QSO. It's true of PSK, JT65, and certainly of JT9.

For me, JT9 is the perfect QRP mode. Try it out!

Thanks, Joe, K1JT

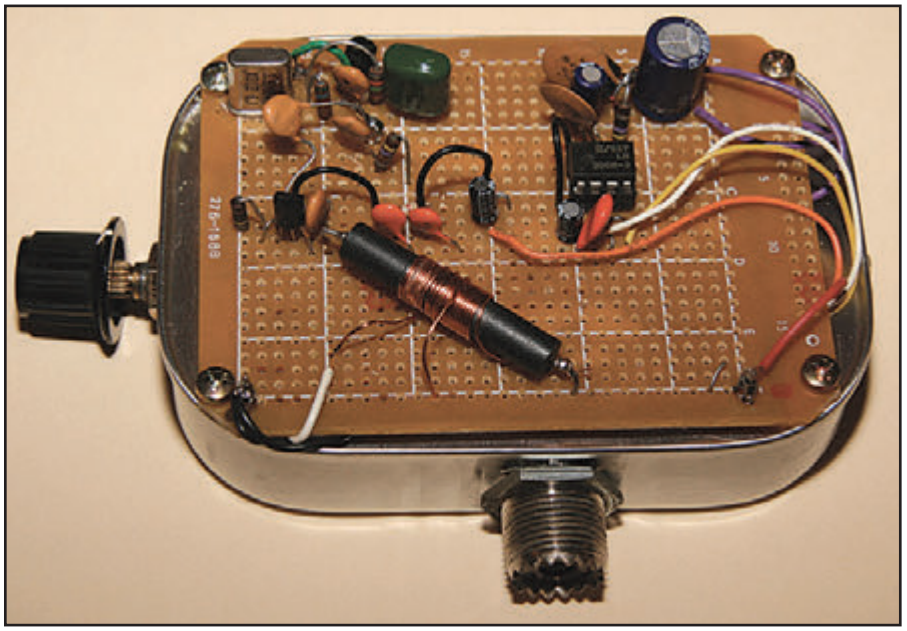
Notes

1. <http://www.pskreporter.info/pskmap. html>
2. Meinberg Network Time Protocol <http://www.meinbergglobal.com/English/sw/ntp.htm>
3. <http://physics.princeton.edu/pulsar/ K1JT/WSJT-X_Users_Guide.pdf>
4. <http://groups.yahoo.com/group/wsjtgroup/messages>

QRP is all about low-power transmitting, often translating into physically small transmitters. Here's a project to build a small receiver—that fits in a sardine tin!—to accompany your tiny transmitter.

Build the 80m Sardine-Ceiver

BY ERIK WESTGARD,* NY9D



NY9D's "Sardine-Ceiver" is a fun-to-build companion receiver for W1FB's classic "Sardine Sender" QRP transmitter.

The concept of an amateur radio "Junk Box" construction project has taken a beating over the last quarter century. Whenever I see one of these, they usually specify one or more components I don't have, even in my extensive inventory. This is after decades of diligent scrap collecting, and a fair bit of deliberate ordering ahead of normal radio construction parts (such as T50-2 toroids) from mail-order suppliers.

The late Doug DeMaw, W1FB, recognized this in his later construction books and articles. He started seeing the supply of variable capacitors and other staples of radio projects disappear from store shelves.Soon after converting the 1978 era all-RadioShack W1FB Sardine Sender to use the newer 100-µh RF chokes, I thought a companionreceiver with only common, locally available parts would be a good project.

I delved into Doug's books and my article archives. The design goals were to have no toroids and to avoid the NE602/612, as it was rare in local stores and in common scrap. I was eyeing a couple of non-working VCRs, which are loaded with RF chokes and other useful components.

I wanted to make the receiver tunable, so I tried out some VVC/varactor-diode tuned VFOs. This effort was very frustrating, and it would not use variable capacitors for tuning or even adjustments. This ruled out most of the available QRP receiver circuit diagrams. The final requirement was to not use dual gate MosFET devices, which were also getting scarce in the stores and scrap piles.

I gave a talk at the Dayton QRPCI meeting called Four Days in May in 2002, and met George Dobbs, G3RJV, a fellow W1FB fan and leading light of the G-QRP Club. He was selling a tiny book called *Minimalist Radio*, which featured a simple mixer based on the MPF102/NTE451 device and had no variable capacitors. The front end was for 40 or 20 meters, and so this book was not an immediate solution, but joined my growing pile of reference articles.

After a few long weekends and several dead-end prototypes, I was not getting far. I even built a Pixie, which was the right idea and the right band. However, the Pixie was prone to BCI (broadcast interference) from a powerful nearby AM station. The events of 9/11 sent me off in the direction of public service and packet radio, and later D-Star, and away from QRP for a while.

Re-awakening a Latent Interest

Recently, I was cleaning out my office and found a W1FB article on an 80-meter-band converter that had a single-ended mixer and an 80-meter front end. The light went on in my head after seven years and I raced for the shop.

The little receiver in George's book was fully modular; you could build it in sections. Therefore, I started with his mixer, and added in a VFO—really a basic crystal oscillator from the selection in Doug's book, *QRP Notebook*. The Colpitts oscillator used a common 2N2222A and a TV colorburst crystal and had no variable capacitors or trans-

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e-mail: <ny9d@arrl.net>

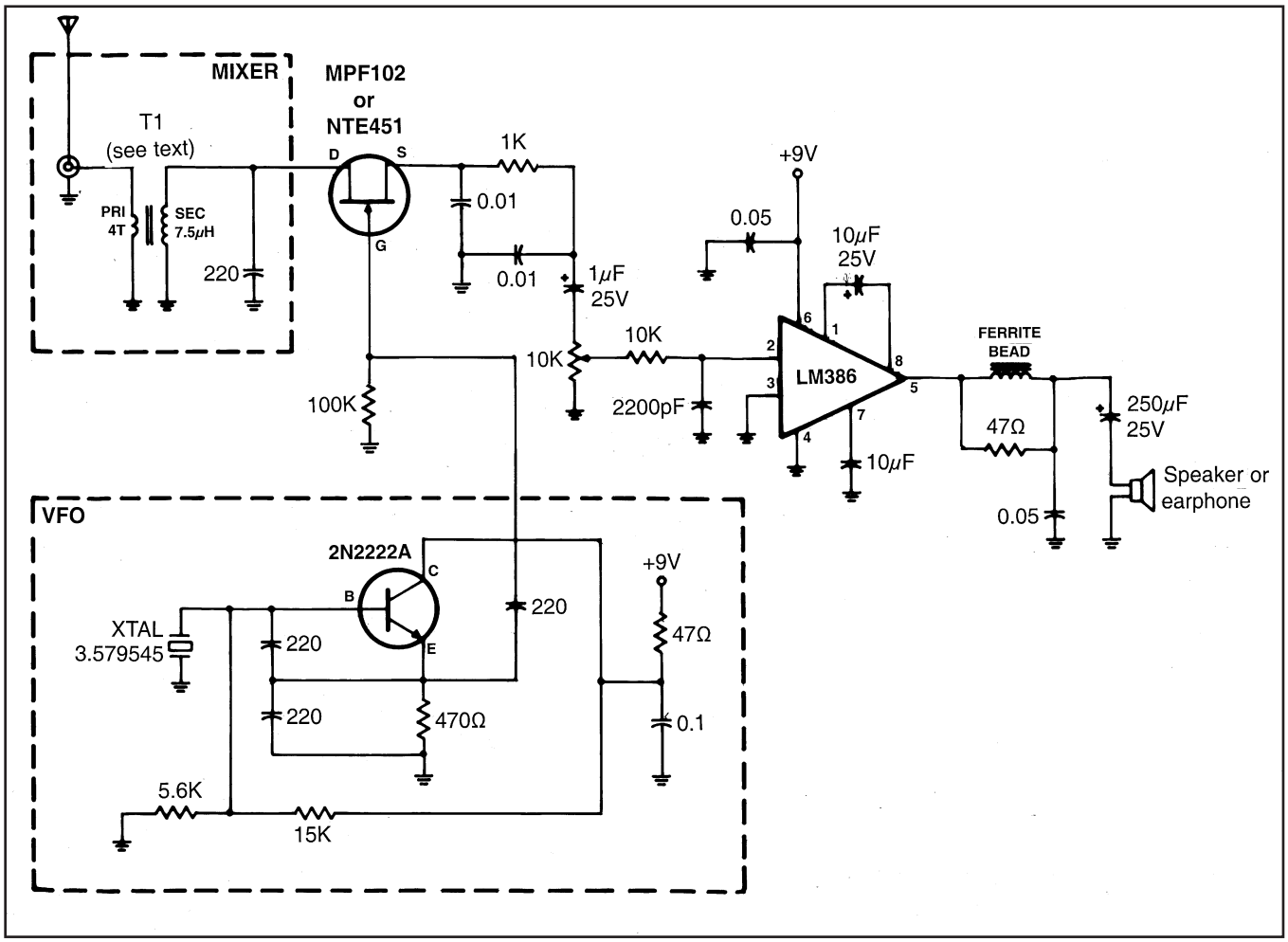

Fig. 1.The SardineCeiver merges a mixer circuit by G3RJV with a W1FB VFO design and a National Semiconductor audio circuit for the LM386. The ferrite bead comes from an old VCR.For T1. Remove all but 19 turns from a RadioShack 100 μH RF choke (273-102), leaving a 7.5 μH secondary; then used some of the removed wire to wind a 4-turn primary over the secondary winding.

formers. The math for the capacitors didn't work out, so taking a tip from Doug, I built a prototype and tried some different caps; 220 pF did the job and it started oscillating. For an audio amp, there was one for an AM radio amp in the National Semiconductor application notes for the LM386. Having had my fill of AM BC RFI, this might be OK.

And for the tuned front end there was Doug's converter article, 7.5 μH as a secondary on a T50-2 toroid and about 220 pF in parallel. Doug had the front end go to the gate on the MPF-102; for the G3RJV design it went to the drain. I said to myself, "Cross your fingers and solder to the drain."

I hooked it up to my RF signal generator and sent in a signal. Yes, it was a receiver, but at the millivolt level not the microvolt level. This was better than the last time, but not good enough. After several frustrating sessions of trying to change values and add stages, one night I took off the signal generator and

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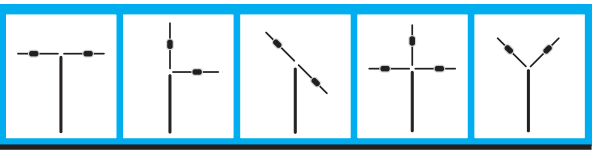

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hooked up a tuned antenna—my G5RV. To my astonishment, clear CW signals from all over the USA were heard. I sat transfixed for a while. George makes the comment (disagreeing with Doug, who does not like single-ended mixers) that it "…works oddly well," which I think fits.

I made up a RadioShack 273-102 100 µH RF choke to match the 7.5 µH secondary (remove all but 19 turns for the secondary and use the leftover wire to make a primary winding over the secondary with four turns). I swapped out the T50-2 toroid transformer for the RadioShack homemade version and it worked the same. I next got out the prototype Sardine Sender transmitter and started calling CQ. After a few hours of that one afternoon in poor conditions, I sent an e-mail to QRP builder AAØZZ, who lived nearby, and asked him to get on the air. Soon I worked him, rusty CW and all. Then it was off to the store for a sardine can.

Most of the parts are available at RadioShack, where the components have moved off the pegboard into drawers in the back. Ceramic capacitors are used for the 220-pF applications. RadioShack has an assortment that might include those, or you can use 100-pF models in parallel. RadioShack used to sell the crystal, but for that now, an old VCR or TV will usually have one. If you don't have one to scrap out (the older the better for fewer surface-mount parts), ask around or check at thrift stores.

Troubleshooting the project consists of building it in stages; maybe start with audio. If it works, you should hear a hiss, and you can tap the input and have it click. The RF bead came out of an old VCR and can be omitted. I think it helps a bit with RF interference. The oscillator should oscillate, and you should be able to hear it in a nearby receiver. The front end and mixer should hear signals (CW, SSB, and even shortwave broadcast stations) if you have it right. Use a good antenna. This little radio is not very sensitive but was nice and clear here. I used a 9V battery, as the LM386 is famous for not liking normal 12–13.6-VDC ham shack power. I got a little local 80-meter SSB interference during the big sweepstakes contest, but no BCI.

The MPF102 is getting harder to find, but NTE451s are still around at places such as Allied Radio, so the mixer design might need to adapt to other locally available semiconductors. I do think the idea of a VFO of some kind is a worthwhile upgrade, as it would be nice to be able to move around the band a little. Of course, too, you can try different circuit modules yourself—such as the audio amplifier—depending on your interests or the parts you have on hand. It is reasonably cheap these days ($20 for one I saw on eBay) to buy a meter to measure inductance, so if you are modifying the RadioShack RF choke (or winding your own toroids) you can check your work.

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Need proof of how far a QRP signal can travel? Try K1JT's WSPR program and see where you're being heard; plus which bands are open to where in real time.

Build Your Own WSPR Beacon

BY YANNICK DEVOS,* XV4Y

The *Weak Signal Propagation Reporter*, WSPR, is a real-time propagation tool invented by Joe Taylor, K1JT. I, for instance, use it as a way to learn the propagation on a band and determine the best time to work a given DX path. Thanks to WSPR, I learned that there is an 80-meter path between the US and Vietnam even in summer, and I got confirmation that the best openings are right after sunrise in North America.

Using WSPR requires a low-power beacon transmitter (the subject of this article) and an Internet-linked computer to let you see where your signal is being heard. If you'd like more background information about WSPR, please refer to G4ILO's excellent article at <http://www.g4ilo.com/wspr. html> or to the presentation by the software's author, K1JT, at <http://www.physics.princeton.edu/pulsar/K1JT/wspr.html>.

Who is This Circuit For?

If you're a homebrewer ham with basic digital and RF skills, this WSPR beacon project (photo A and fig. 1) is for you. There is no need for laboratory measurement equipment such as an oscilloscope. Everything should run straight from the beginning. However, some care and patience may be required, because the parts used will not accept any overvoltage or polarity inversion.

Component parts for this project (see photo B) are readily available (see below) or you can build it from a kit with a PCB and all the parts included. The kit I recommend is the Ultimate 3 QRSS Beacon from Hans Summers, GØUPL <www.hanssummers.com>. *(QRSS refers to very slow speed CW, as often used on the VLF, or very low frequency, bands.—ed.)*

I used to sell what I called "semi-kits" for this circuit (photo C), but now Hans has carefully refined its design and his kit offers more value for the money. Some of our choices are different (for example, my source code is freely available and easily customizable), but you can make your own decision.

Design Choices

I first started to work on this project two years ago. KD1JV (father of the ATS/MTR transceivers) introduced me to the MSP430 family of micro-controllers and I quickly wrote a small WSPR keyer to use with my QRSS beacon (original beacon kit from Hans, GØUPL). The main advantage of the MSP430G2553, as it is provided with the LaunchPad board from Texas Instruments, is the ability to drive a 32.768-kHz crystal like you find in clocks. Then with a few lines of code you can have a fairly accurate *Real-Time Clock (RTC).*

Photo A. Screen of the author's WSPR beacon while transmitting a Weak Signal Propagation Reporter *signal.*

Why do we need an RTC for a beacon? Because WSPR is very time sensitive. In order to recover the data through the QRM, it uses a known time frame to guess the missing digital symbols from the other good ones. If you're off more than 2 seconds from the other stations, you will see that the successful decoding rate will drop significantly. Most WSPR users have the K1JT software running on a computer with the time set via the *Network Time Protocol* (http://www.ntp.org/). The first WSPR standalone beacon designs all used a GPS-controlled clock. Two years ago, dedicated GPS modules were expensive (and they still are somewhat), so that was a no-go for me.

The MSP430 Launchpad enables me to have a WSPR beacon running for up to three days with accurate timing. For longer durations, it was drifting too much from real time, but the only thing I had to do was manually set it to second zero of an even minute, using my smartphone's internal GPS with an Android application such as "Atomic Clock."

Parts Providers for the LaunchPad, Display and DDS Module

Well, for the MSP430 LaunchPad evaluation board, you should go straight to Texas Instruments (http://www.ti.com/tool/msp-exp430g2). It's available there for less than $10 with all the required parts and even more.

*e-mail: <xv4y@nature-mekong.com>

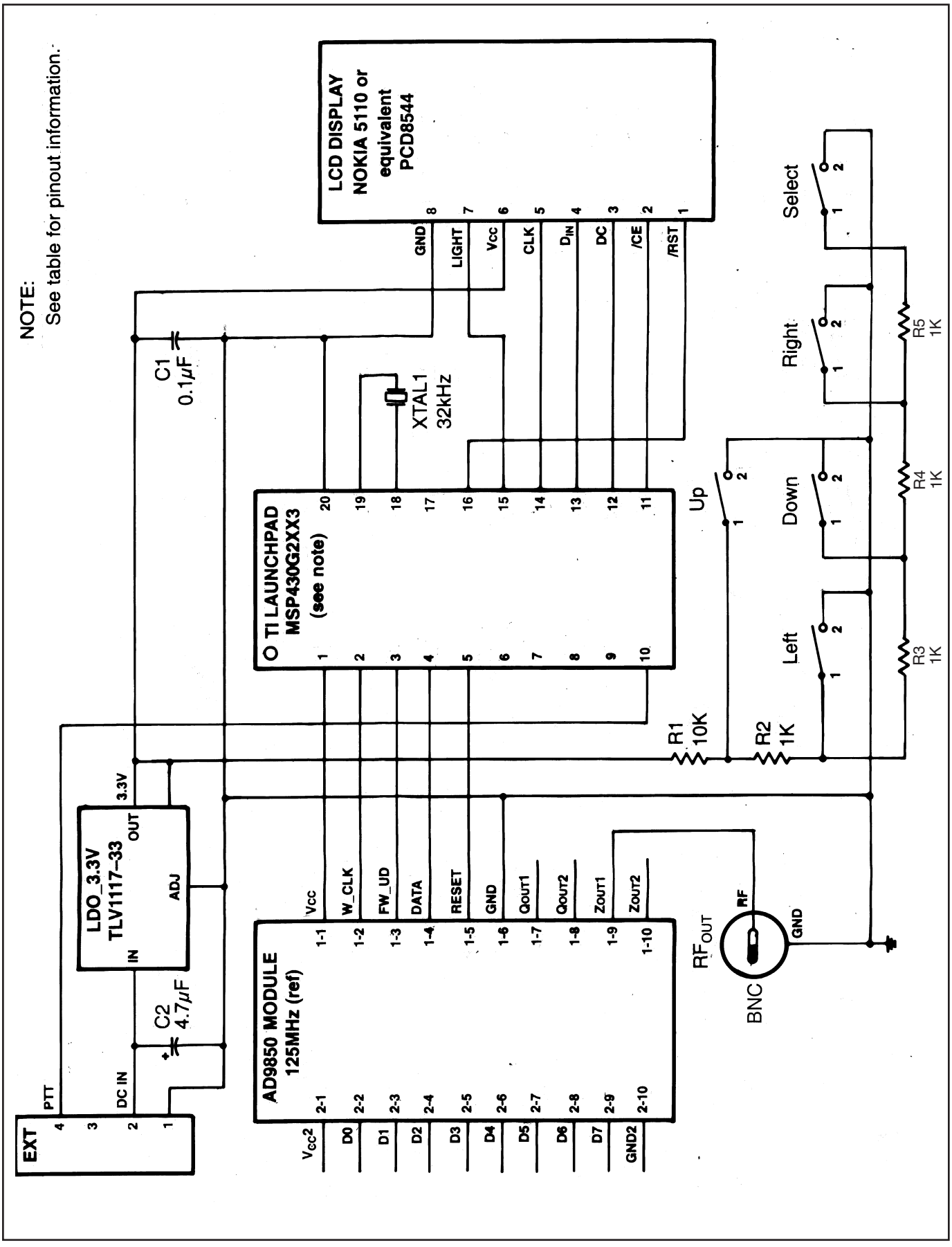

Fig. 1. Schematic of the WSPR beacon transmitter. See Table I for Launchpad chip pinout information.

Pin #	Function
1	DVCC
2	P1.0/TA0CLK/ACLD/A0*/CA0
3	P1.1/TA0.0/UCA0RXD/UCA0SOMI/A1*/CA1
4	P1.2/TA0.1/UCA0TXD/PUCA0SIM0/A2*/CA2
5	P1.3/V_{REF-}*/VEREF-*/
6	P1.4/V_{REF+}*/VEREF+*/TCK/ . . .
7	P1.5/TA0.0/UCB0CLK/TMS/ . . .
8	P2.0
9	P2.1
10	P2.2
11	P2.3
12	P2.4
13	P2.5
14	P1.6/TA0.1/TDI/TCLK/ . . .
15	P1.7/CAOUT/TDO/TDI/ . . .
16	_RST/NMI/SBWTDIO
17	TEST/SBWTCK
18	XOUT/P2.7
19	XIN/P2.6/TA0.1
20	DVSS

Table I. MSP430G2XX3 (TI Launchpad) Pinouts. See schematic, fig. 1, for additional connection information.

The DDS module, Nokia 5110-like display, and small parts (switches, cables) are readily available from Chinese sellers on eBay or many suppliers in the United States. My provider of choice is called "Chip_Partner" <http://www.ebay.com/sch/chip_partner/m.html>, but there are many. I also have had good experience with a website named IC Station <http://www.icstation.com/>. I have no relationship with either apart from the fact that they are my usual providers. Prices have decreased tremendously and should be around $5 for the DDS module and $3 for the LCD.

My own experience led me to buy the AD9850 modules with only two rows of pins and a big blue trimmer. I prefer the ones with a full-size canned oscillator (not the SMD version), as they usually have better frequency stability. An AD9850 with 125-MHz reference clock will allow you to generate signals between a few Hz and 30 MHz. (Pin 1 of each row is identified by a square.)

WSPR and QRSS are very demanding in terms of signal purity and frequency stability. I had only bad experiences with the AD9851 modules and the "three rows" modules. Even with the AD9850 two-rows modules, it is not unusual to have 10% of the parts that are a bit on the edge regarding specifications. It is not unwise to buy two of them, as they are cheap. Keep the best one for WSPR; the other one will be good as a VFO for a small receiver.

A last note about the LCD: It is not required in the WSPR beacon source code I am providing. However, given its price, I would not hesitate long, because without it you will lose a lot of flexibility.

Source Code for the Firmware

On my own website I made available three source codes for firmware running on this circuit. This programs must be compiled and uploaded to the LaunchPad's MSP430G2553 using

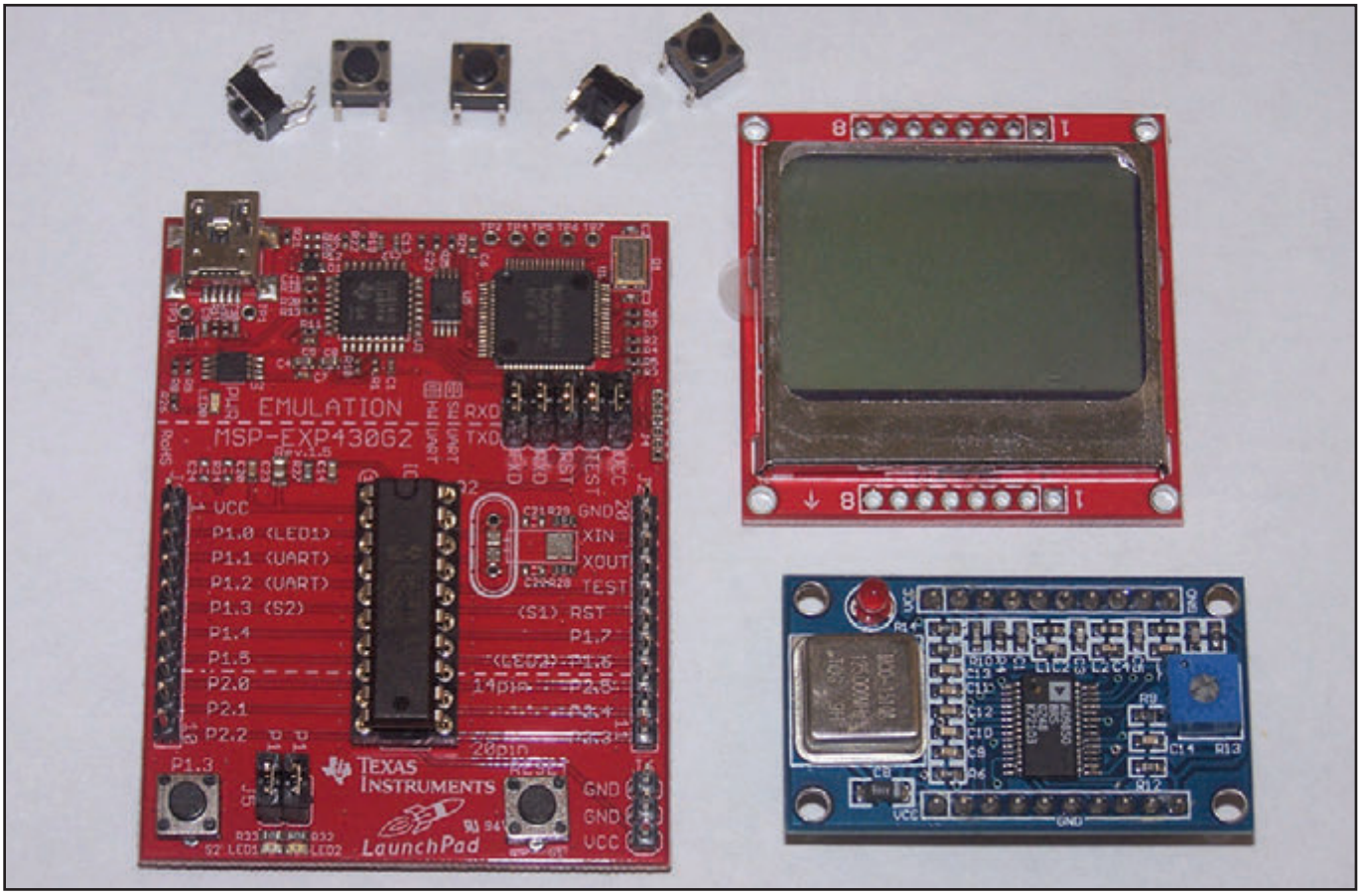

Photo B. Basic parts needed for the beacon include the Texas Instruments Launchpad (left), LCD display (upper right), AD9850 module (lower right), and several switches.

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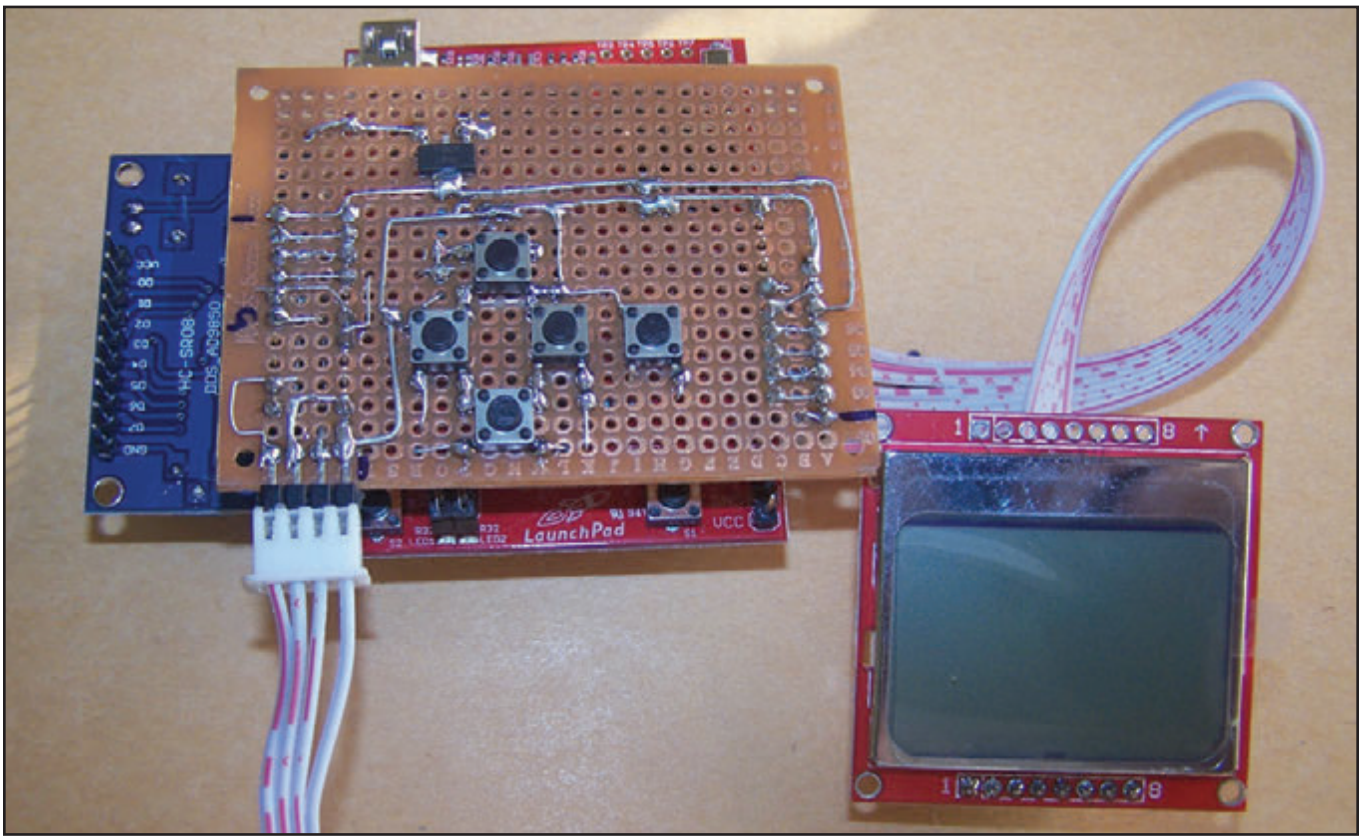

Photo C. The author's semi-kit for the WSPR beacon that he used to sell. He now recommends purchasing a kit from GØUPL (www.hanssummers.com).

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the Energia IDE. All details are on this web page: <http://xv4y.radioclub.asia/cq-magazine-wspr-beacon/>.

With this circuit and compiling/uploading the firmware you can build:

- A full-featured WSPR beacon with runtime parameters to select the callsign, grid locator, band, frequency, and WSPR version (standard WSPR-2 or WSPR-15 for VLF).
- A QRSS beacon with menu selection for frequency, keying width, band cycling.
- A very simple DDS controller with VFO A/B and band selection to build a signal generator or a small transceiver local oscillator (LO).

In order to compile the given codes you will need to download and install them in the Energia libraries directory. Two are written by me (RTC Clock and AD9850 control) and one is a third-party library (Nokia-5110 display control). All are available on the previously indicated web page as well as the user's manual for the firmware explaining the settings and beacon behavior.

How to Interact with the Micro-controller

On my kits I used to install five push-buttons, four as up/down and right/left and one for validating. The WSPR firmware uses only three: up, down, and validate. You are free to only solder three buttons, but you may have to change the source code and will not be able to use the VFO software, which needs five buttons. For the price of two buttons and two 1k-Ω resistors, I recommend that you install all five buttons.

If you're using the same resistor values I am (10k and 1k), then the original source code should work. If you're using different values (or low-precision parts), you may need to adjust some numbers in the source code for mapping with the internal ADC measurements. In that case, the simple "AnalogInSerialOut" example code provided with Energia will give you the right values for each button-press.

There is also a PTT signal on Pin 2 of Port 2 of the MSP430. It is high when transmitting and low when idling. While idling, the DDS is put in "power saving" mode, turning off its output. If you run a class C or D/E amplifier, this will be enough to save power. If you want to do trickier things, feel free to use the PTT signal.

Which Voltage for DC Input?

Using a *Low Drop-Out* linear regulator such as the LVT1117-33, input voltage can be between 5V and 10V. If you want to power it from more than 10V (if you have a 12V rail from your PA), just add two or three 1N4007 diodes in series with the regulator input.

For simplicity, I only use one 3.3V Vcc rail for the MSP430G2553, the LCD screen, and the DDS module. The AD9850 is also happy at 5V and will even put out a little bit more RF power at this voltage. When using AD9851 modules (which I don't recommend), this is necessary because the 160-MHz clock may require them to be powered from 5V (even if the module's specifications states the opposite) in order to have reliable functioning.

You can add a separate 5V rail solely for powering the DDS, but take care not to mix both 5V and 3.3V rails because you can easily fry the MSP430 MCU.

At 3.3V, the circuit (LaunchPad + AD9850 + LCD) draws less than 500mA. If you run the AD9850 at 5V, it will draw more current.

Just as a comment, you can also power the circuit through the USB port of the LaunchPad. However, I don't recommend it except for testing and debugging. While using it "in pro-

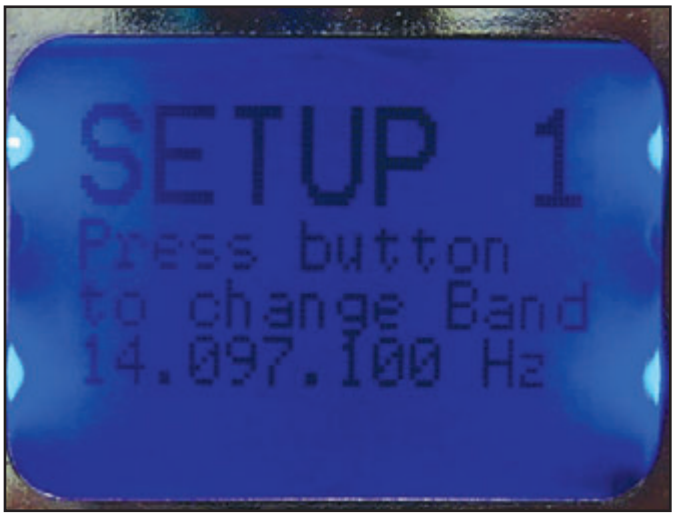

Photo D. Everything should work first time, every time. This is the setup screen you should see on initial power-up.

duction," I recommend that you open the five jumpers connecting the two parts of the LaunchPad and only power the LaunchPad from an outside source. Of course, never connect both power sources at the same time!

Building Advice

There are no special recommendations and everything should work from the first try (photo D). Sometimes, with some screens, when powering on the whole circuit, the LCD does not initialize itself well (maybe it is not ready while receiving the reset commands). If this happens, try to do a reset using the dedicated button on the LaunchPad board (putting the RESET signal to low).

Try to keep your connections short, but frankly speaking, I have been running "hairy" prototypes with long connections and everything still works well.

If you want to see pictures of how I actually built my own beacon, you can have a look at my blog. Most pages are in French, but an automatic translation is provided so you should understand what I'm talking about. Here is the latest article concerning this project: <http://xv4y.radioclub.asia/2013/06/23/montage-dun-kit-filtre-passe-bas-universal-de-kits-and-parts/>

Should You Use a Power Amplifier?

The DDS board output's spectral purity is rather good since it has a fairly high clock reference and 70-MHz low-pass filter. However, if you plan to use this kit for frequencies below 2 MHz or above 25 MHz with a high-gain amplifier, more filtering may be required.

As is, the DDS output power level is around 4 dBm. This is all right for a VFO or a signal generator, but a bit low for a beacon, even with WSPR. One watt is a good target level for such usage, as WSPR is considered at least 10 dB more "sensitive" than CW.

On HF, several class C power amplifier (PA) designs can provide output of about 5W for 10mW input. Kits and Parts (photo E) and the DL-QRP Group offer high-quality kits for this purpose. The latter is more expensive but may be used on higher frequencies. The ZOUT1 output of the DDS board has 50-ohm impedance, so you should be able to hook it directly to either of those PAs.

Using the comparator output (QOUT1) and adjusting the blue potentiometer to have a 50% duty-cycle, you can also feed the RF to a class E power amplifier as Hans, GØUPL, does on his Ultimate 2 and 3 QRSS beacon kits. You may grab its design or the one from Steve, KD1JV. They use digital parts such as the 74HC00 as driver and several Bs170s (or 2N7000s) in parallels for the final.

A Few Final Words

I hope you enjoyed reading this article and that you will have fun building your own project. All my source code is available under GNU GPL License for non-commercial use and may be modified to better fit your needs. Feel free to leave comments on my blog page dedicated to this article and I will try to answer your questions as soon as possible.

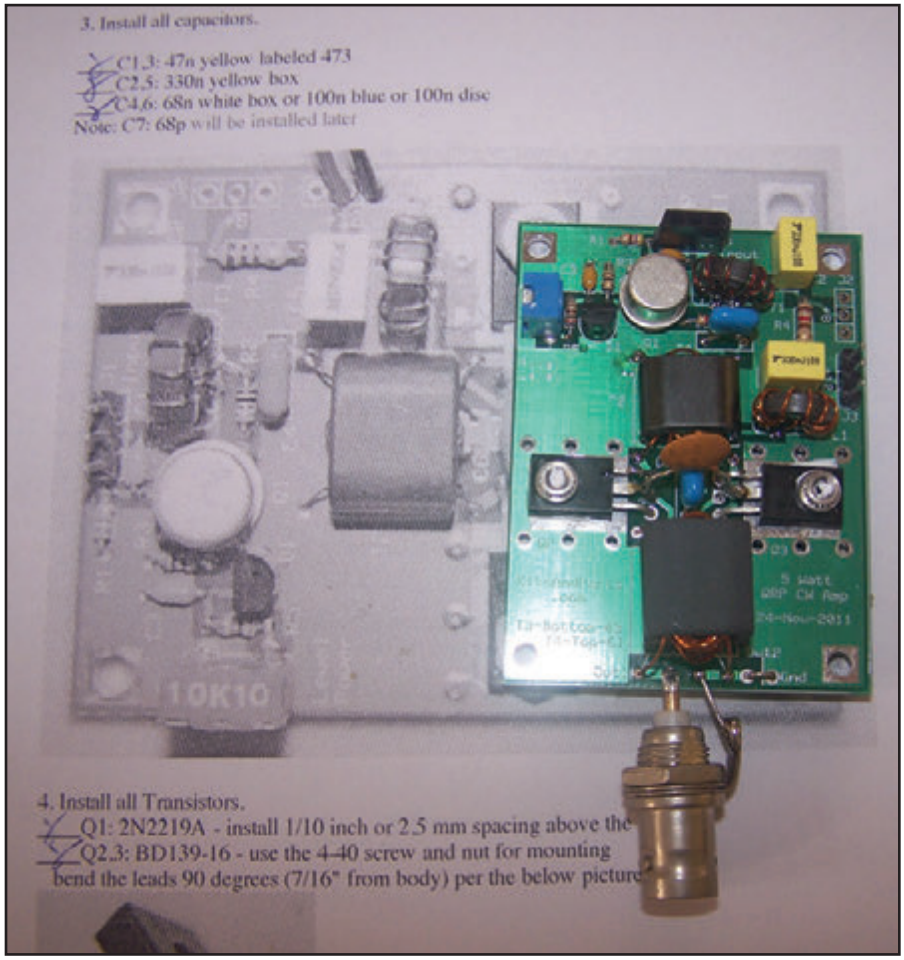

Photo E. Kits and Parts (www.kitsandparts.com) offers a class C power amplifier that may be used with the beacon transmitter. Similar amps are available from other sources as well.

Using a wireless digital network may not seem like traditional QRP, but transmitting with low power is at the core of how these networks operate. Here's a basic guide to setting up your own.

Setting Up a Mesh Network Node

BY J. GORDON BEATTIE, JR.,* W2TTT

You may have read about "mesh nodes" or "mesh networks" and wondered how these amateur radio nodes get configured to produce a functional network. Amateur radio mesh nodes based on the *Austin Broadband-Hamnet (HSMM-MESH) Group's* wonderful software take care of the identification requirements while also creating functional networks that can be used for both routine and emergency communications.[1] With this software, readily available Linksys WRT-54G/GS/GL routers (photo A) can be transformed into mesh network nodes that provide fast and robust self-organizing networks through which we can pass text, data files, video, and even voice traffic.

There is an inherent simplicity to the Linksys implementation that is quite attractive to the new user of this technology and it is strongly recommended as an entry-level system. In this article, we will be discussing the setup and configuration of the Linksys-based HSMM-MESH/Broadband-Hamnet nodes.

Getting Started

You will need to obtain a Linksys router, and that requires some care to get a model and version that are compatible with the mesh software. The compatible models use the Linux operating system. You can find the model and version information on the bottom of the router in the upper left corner of the label right under the Linksys logo as shown in photo B. See Table I for a listing of compatible models and required software. If you find a router on the Internet, it s likely to be pre-loaded with an aftermarket software build called "DD-WRT" which provides some additional conventional router functions. If you find such a unit at an attractive price, don't be worried about compatibility as long as the version is listed in Table I. Of course, the best unit is one that has the correct version, powers up, and is *free*, so check garage sales and local thrift stores. For reference, the WRT-54GS versions 1 through 3 are the best, because they provide the most memory for embedded applications to be resident on the mesh node.

Photo A. Three different models of Linksys routers (from top, WRT54GL, WRT54GS and WRT54G) that are ideal for use as nodes in Broadband/Hamnet (HSMM-MESH) networks.

*e-mail: <W2TTT@att.net>

Model	Version/Notes	Software Version
WRT-54	V1-4 (V1 is 5V only)	bbhn-1.0.0-wrt54g
WRT-54GS	V1-3 Most desirable has large memory capacity	bbhn-1.0.0-wrt54gs
WRT-54GS	V4	bbhn-1.0.0-wrt54gs_v4
WRT-54GL	V1.0–1.1	bbhn-1.0.0-wrt54g

Table I. Mesh network-compatible Linksys WRT-54/WRT54GS/WRT-54GL models.

Once you have your router, plug it in and make sure that it cycles up in a few seconds. If successful, the "Power" and "WLAN" LEDs will be lit. Next, connect your computer's Ethernet port to any one of the router's LAN ports, labeled 1–4, then open up a web browser and enter the IP address <http://192.168.1.1> to log on to the router's administrative console. If the router is in its default mode and is running Linksys software, then you will need to enter nothing in the user name field and "admin" in the password field. If you are repurposing a router with DD-WRT, then you will use the logon of "root" and "admin" as your password. One of these should let you into the router's administrative console.

Troubleshooting Router Access

If you have trouble accessing the router's logon screen, then you may need to get a new IP address from this router. For Windows® PCs, go to a system command prompt[2] and type "IPCONFIG/RELEASE <cr>" and then "IPCONFIG/RENEW<cr>" (see fig. 1). You should then be able to access the router and log on. Apple and Linux-based systems have similar commands.

If you have trouble with the logon and password because you either don't know them or forgot them, then power on the router while holding the reset button down with a pen for about 30 seconds. This will clear the whole configuration and should restore the router to default settings.

Obtaining and Loading the MESH Software

There are several different files to use with your Linksys router depending on its model and version. You will find the correct file name for your hardware in Table I. Download the needed file from either the Broadband-Hamnet web site <http://www.hsmm-mesh.org/software-download.html> or from the file section of the "ComTechGroup" Yahoo Group <http://groups.yahoo.com/group/ComTechGroup/>, which you can easily join by simply requesting access.

Once the file is saved on your computer and you have logged into the router, select the "Administration" section tab and then select the "Upgrade" button. You will be prompted for a filename. Browse over to the correct folder and select the file. *At NO TIME during this phase of the process should you unplug the router from the PC.* Once completed, you will note that the router reboots and is now an unconfigured mesh node. Your router will take a bit longer to boot the mesh software than it did the default Linksys or other router software. The "Power" LED will begin to flash, then the "DMZ" LED will come on. Once the "DMZ" LED goes off and the "WLAN" LED starts to flash, you have successfully loaded and rebooted your mesh node! You will see that the header of the page has your call sign and identifier on top, along with a page of statistical elements. *Congratulate yourself!*

Basic ASIC MESH Node Configuration

You're making progress, but you have more to do. First, you will probably need to repeat the process of releasing the old IP address and obtaining a new one as you did when initially accessing the console. (These addresses are generated by the router itself. See fig. 1.) Once you have the new IP address, then enter "http://localnode:8080 <cr>" to see the main mesh node console as shown in fig. 2. Click on the "Setup" button and enter a logon of "root" and "mesh" as your password.

You will then be presented with a basic configuration screen as shown in fig. 3. You will need to enter a new value for the "Node Name." It should include your call sign with an extension that makes the node unique with respect to its ID. Some folks use functional extensions with serial numbers, while other put in locations. There are no hard and

Photo B. The bottom of a Linksys router will show you its model number and version information.

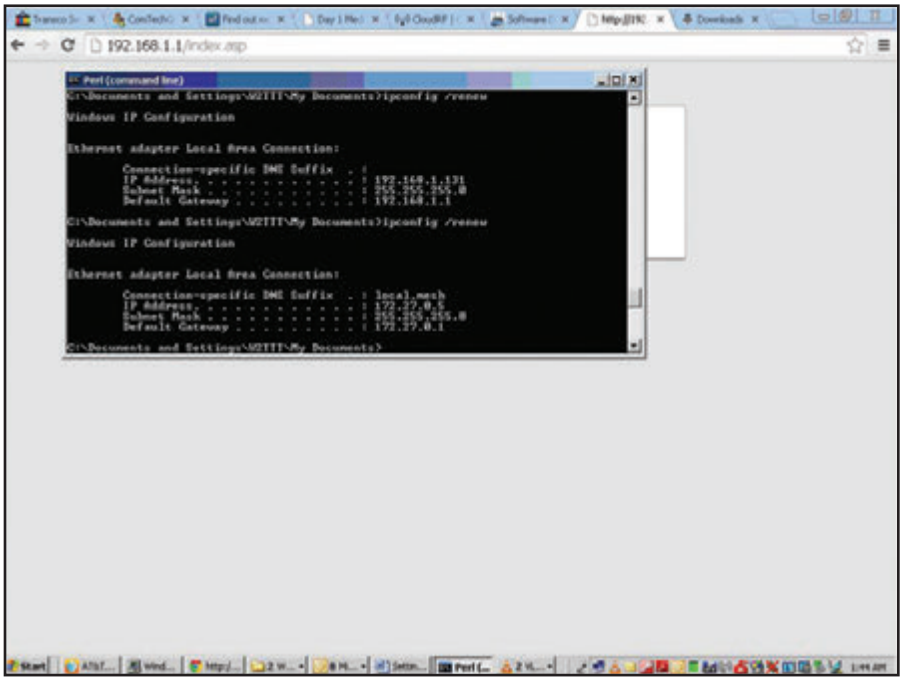

Fig. 1. Windows IPCONFIG example. Note the connection information of "local.mesh" in the lower part of the screen.

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fast rules, but make sure that your call sign is at the beginning. You will also need to enter a new node password in two places. Make sure that you remember what you entered. You might even want to put a label on the bottom of the node with that password as long as you can maintain physical security, but in any case, *write it down—somewhere*!

Finally, it is strongly recommended that you move the node off of channel 1, 6, or 11, as they are the most popular ones in use for normal WiFi operations. We usually pick channel 3, but your local RF ecosystem may drive other choices. (*Note:* WiFi channels 1–6 are shared with the 2.3-GHz amateur band, so any planned activities under FCC Part 97

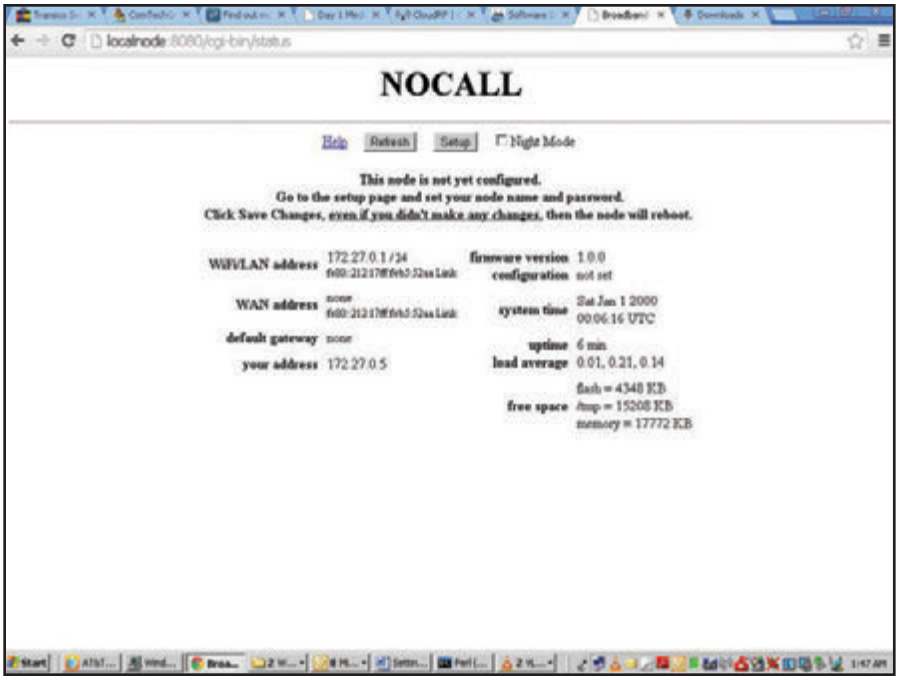


Fig. 2. This is the screen you will see after loading the mesh software but before configuring the node. You now need to do setup.

Fig. 3. The configuration screen will guide you through the steps needed for correct setup.

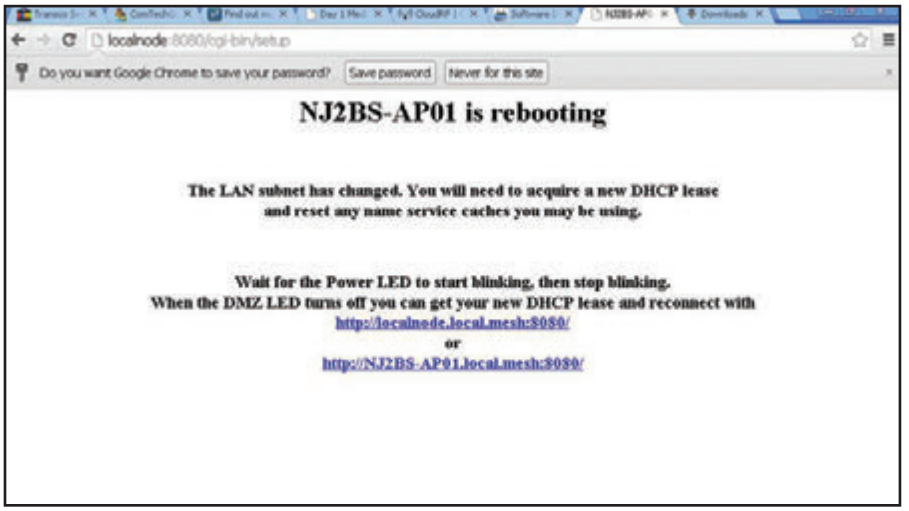


Fig. 4. Once you've entered all the necessary information, you'll get a screen that looks like this while the router is configuring and rebooting itself.

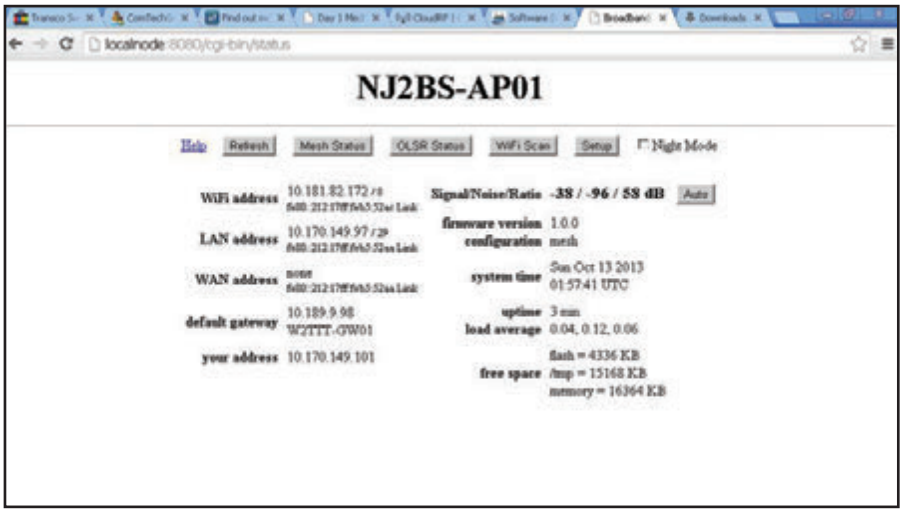


Fig. 5. Voila! Your mesh node is configured and ready to start communicating!

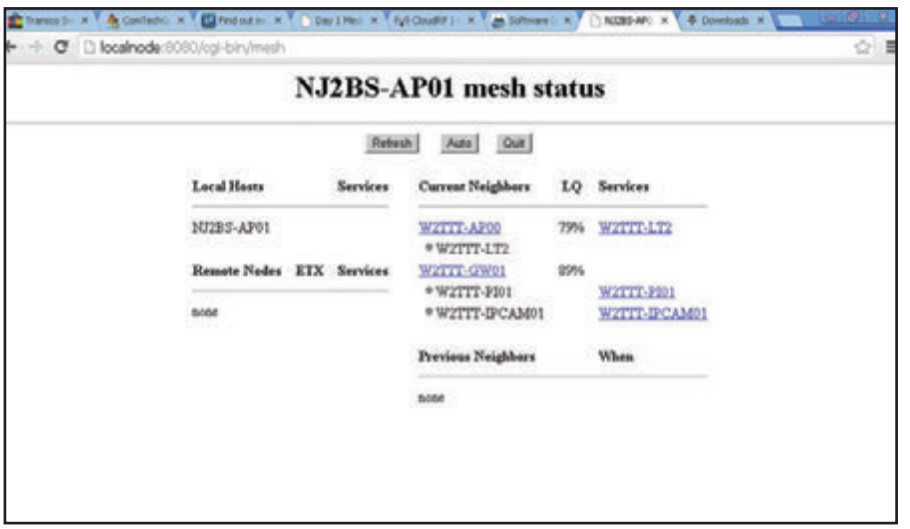
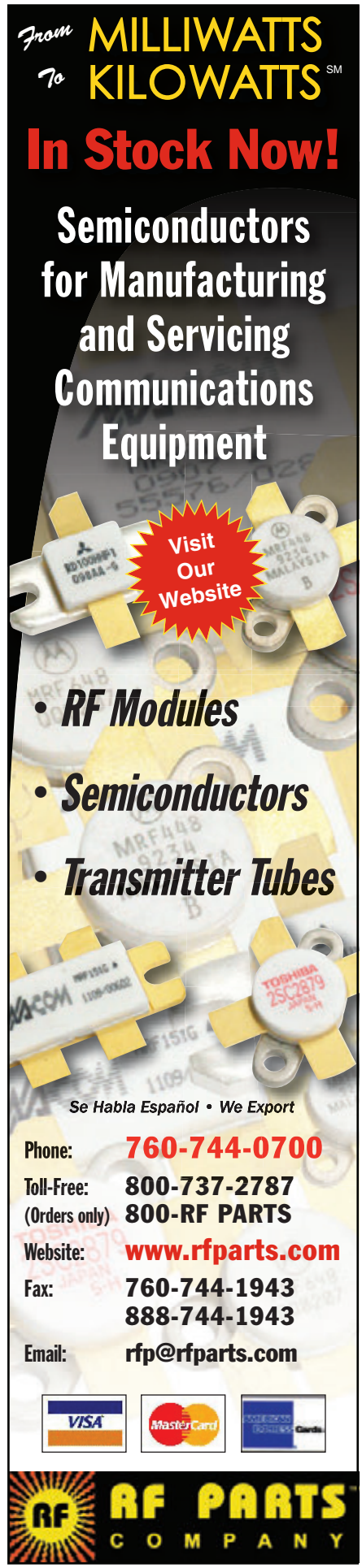


Fig. 6. The mesh network status screen shows you what other nodes are accessible to you as well as what features, or services, they offer.



rules—which allow more power and antenna flexibility than Part 15 rules—must be conducted on one of these channels.)

One interesting thing to note is that if two mesh nodes are on different channels, they will search out other nodes on other channels, but at a cost in network performance. Once it is all entered, press the "Save Changes" button and the node will save your entries and reboot. During that reboot period you will see a screen as shown in fig. 4. Once it comes up as described previously, you will need to once again ensure that the computer gets a new IP address. Once that is done, enter "http://localnode:8080 <cr>" in your web browser, and you will be presented with a screen as shown in fig. 5. You now have a functional configured mesh node that should be ready to communicate. *Congratulate yourself again!*

Fig. 7. The Foscam IP camera lets the user select different modes based on which one works best on your browser. The camera is accessed and configured over the mesh network.

Fig. 8. Example of what you might see from the IP camera once you've selected a mode on which to view its photos.

Is There Anybody Out There?

OK, I borrowed the title line from Pink Floyd, but there really is a need to put up more than one node and find out if they can communicate with each other. In order to see neighboring nodes, select the "Mesh Status" button. If you have two configured nodes or have a nearby node, you will see something like what's shown in fig. 6.

As you can see, there are two neighboring mesh nodes and each has some defined "Services." You can connect to these services if they are defined and are using a compatible device. For example, W2TTT-LT2 is a Telnet server that allows for keyboard-to-keyboard connections, while W2TTT-PI01 is a Raspberry Pi computer running an Internet Relay Chat (IRC) server. The W2TTT-IPCAM01 is an IP camera. Connecting to this Foscam brand camera is as easy as clicking on the camera link from your web browser and following the camera server connection sequence (fig. 7). Along the way you will be prompted for a camera logon and password, and then you will be provided with a choice of camera server modes. Try each one with your browser to determine which option will give you success (fig. 8). ActiveX Mode provides full recording and snapshot capabilities which are not shown here as that plug-in was not installed into the Chrome browser that was used for the screen shots in figs. 7 and 8.

Just Do It!

We all have become interested in a topic that we've read about, gotten really fired up, and then came the practical reality of making it happen. You know the type of issues that come up…"I'm distracted," "I don't know how to find or load the new software," "What happens if I do it wrong?" etc. This article is an attempt to help you conquer those issues, and *get on the air*! I hope that you found it helpful and interesting.

Notes

1. See Beattie, "HSMM-MESH Networking in EmComm Applications," *CQ* October 2013, p. 26

2. To access the system command prompt in Windows, click on the "Start" or Windows logo button, then type "CMD" in the search bar and click on "CMD.EXE" when it comes up.

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When a simple wire antenna is either all you want or all you can use for your low-power station, consider an end-fed half-wave (EFHW) with KI6SN's combo tuner and SWR bridge.

A QRPer's 40- through 17-Meter EFHW Antenna Coupler with Super-Sensitive SWR Bridge

BY RICHARD FISHER,* KI6SN

Some radio amateurs are QRPers by choice, leaning on efficiencies of their radios, antennas, and operating skills for the thrill of logging contacts up the block and around the world with transmit power of 5 watts or less. Others are QRPers by necessity. They live in cramped quarters where the guy next door's TV, AM-FM radio, and WiFi may be only a thin wall away. Even a 100-watt transceiver will cream them, let alone a kilowatt. Besides, the apartment manager or homeowners association may say outdoor antennas are *verboten.* Therefore, the op hunkers down at 5-watts output or less and lets necessity drive an aerial solution.

For all sorts of good reasons, the end-fed half-wave (EFHW) wire antenna with a counterpoise has been a popular means to that end—for good reasons. They are:

- Simple, inexpensive systems to make
- Suffer no feed-line loss, because *there is no feed line*
- Can be selected for the band or bands of operation based on what your real estate will bear, whether indoors or outdoors
- Need only one support to be erected

*(**IN DEPTH:** What is a counterpoise? Visit:* <http://bit.ly/19mwy9t>.*—KI6SN.)*

Of course, there is no free lunch. The catch here is that your transceiver is looking at an EFHW antenna impedance running in the range of 1,800 to 5,000 ohms—certainly not the 50 ohms for which your radio was designed.

Fortunately, there is a very simple solution. It is one that can be achieved through the work of your own hands:

Build an EFHW antenna coupler, and throw in a *super-sensitive* absorptive SWR bridge to boot (photo A)! We'll call it the *40- through 17-meter EFHW Antenna Coupler with Super-Sensitive SWR Bridge,* which grew from the ideas of two outstanding radio amateurs, who we'll get to in a minute.

First, a Word from Our Sponsor

There are, no doubt, some skeptics getting ready to turn the page. *"An end-fed antenna with a cheesy coupler at QRP levels? What kind of success can you expect with a lame set-up like that?"*

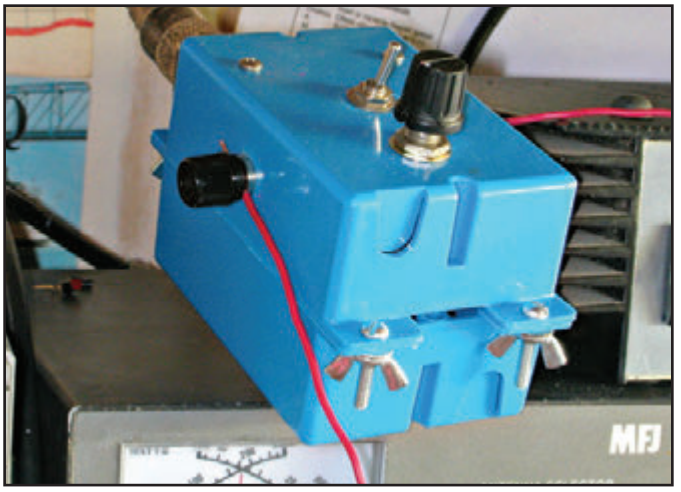

Photo A. With just a handful of parts and a couple of hours at the workbench, this 40- through 17-Meter EFHW Antenna Coupler with Super-Sensitive SWR Bridge *can be your ticket to getting on multiple bands with just a wire antenna and counterpoise. (Photos by KI6SN)*

Believe me, and others who swear by (*not at*) EFHW antennas: They work, even at 5-watts RF output. They can work *really well.*

Here's a KI6SN testimonial. After finishing this coupler-SWR bridge project and:

- Drooping the transceiver-end of a 66-foot, 40-meter half-wave wire from my operating bench through the rafters to the front of the garage ...
- Folding the wire back on itself across the roof to a PVC pole held to a pipe with a bungee cord ...
- And turning the wire 90 degrees horizontally to be tied off at the chimney ...

... We laid a 33-foot long counterpoise on the garage floor out onto the driveway and got on the air.

It was the Sunday morning of the CW leg of the CQ World-Wide DX Contest last November. I tuned the Kenwood TS-140S to 40 meters and peaked the SWR bridge for the best

*Contributing Editor, Public Service Editor, CQ
e-mail: <ki6sn@cq-amateur-radio.com>

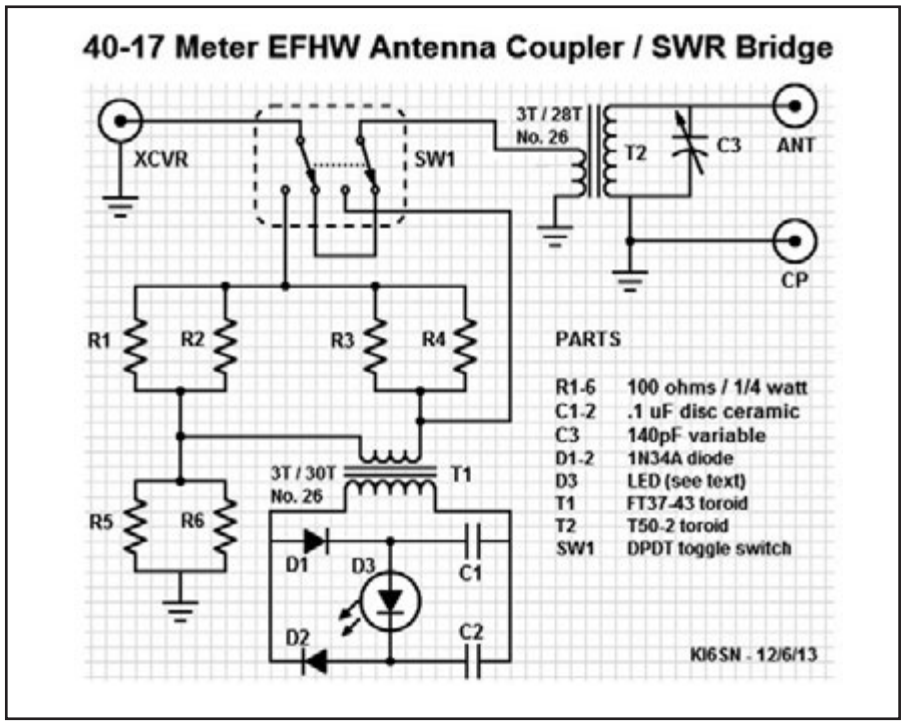

Fig. 1. Note that one wire crossing another without a dot means there is no connection between the two. For example, look at SW1. There are four solder points, left to right, shown at the bottom of the DPDT. Points 2 and 4 are connected, but do not connect to the wire going down and right from point 3. (Courtesy of KI6SN)

match. Running a full "QRP spoonful" (5 watts), I rattled off 14 Japanese stations, a Russian near the border of Mongolia, a Hawaiian, and several Canadians—all in the course of an hour around southern California local sunrise.

Later in the day on 17 meters —with an 18-MHz EFHW dangling about five feet off the garage floor—Randy Dominicak, WA6RKK, just south of the Canadian border in Rathdrum, Idaho, gave me a 5-7 on SSB. Again, I was running 5 watts. This time, though, with a completely "indoor" EFHW wire and counterpoise lying on the garage floor.

Yes, it's a tribute to WA6RKK's and all the DX stations' great operating skills, antennas, and receivers. By the same token, though, it proved an EFHW antenna, for all its simplicity, can be an amazing radiator.

Great Minds ...

When setting out to build a coupling unit capable of multi-band operation, we turned for inspiration to two experts in the amateur radio antenna game:

• Steve Yates, AA5TB, of Fort Worth, Texas, generally considered the world's

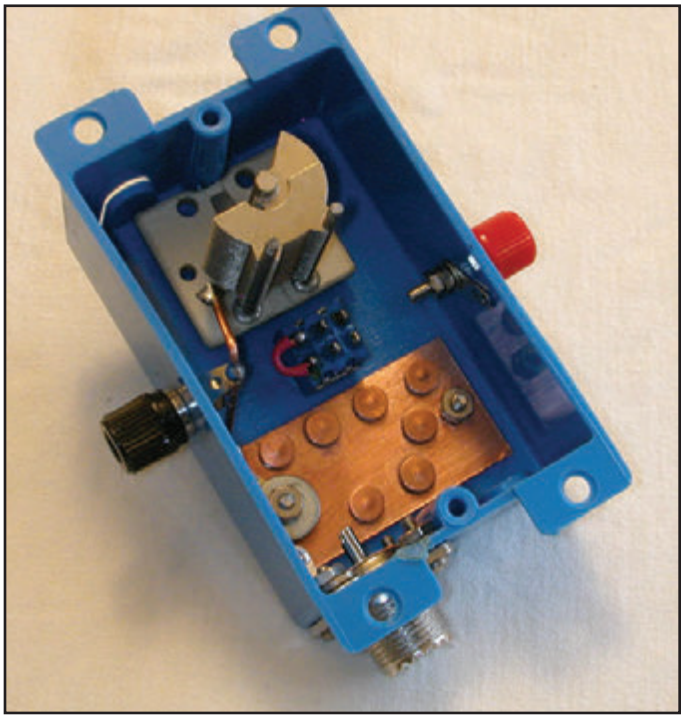

Photo B. The 140-pF air-variable capacitor, where the coupler circuitry will be positioned, is at the top of the box. The SWR PC board, with seven Manhattan pads attached, is at the bottom.

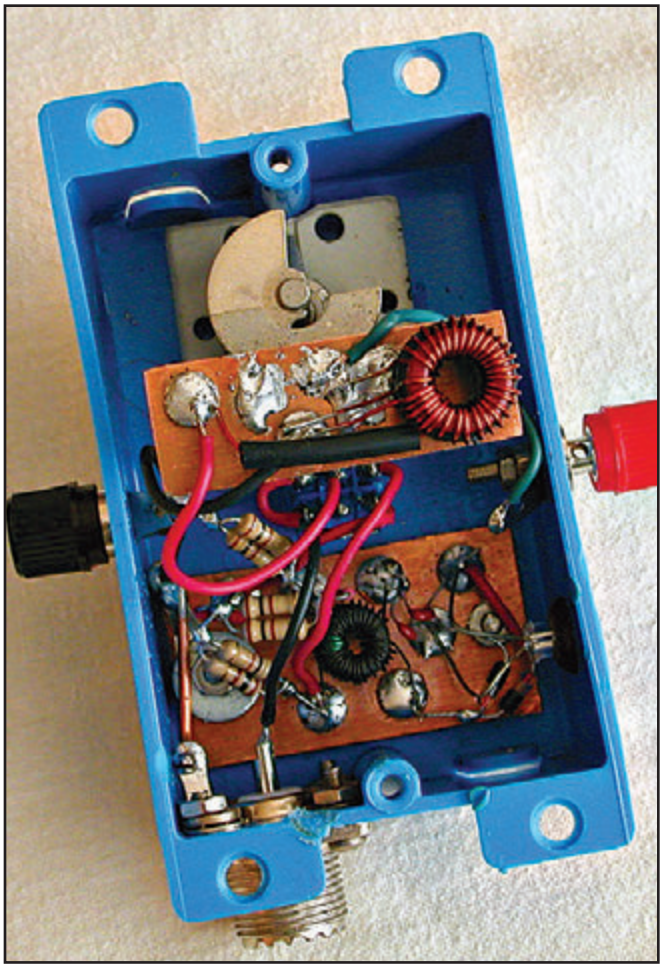

Photo C. It doesn't take long for 8 cubic-inches to be filled with antenna coupling and SWR-sensing components.

leading authority on EFHW antenna coupler design and operation. Visit his website for a textbook's worth of EFHW information at <http://bit.ly/ 1eMGWeL>.

• Glenn Leinweber, VE3DNL, of Hamilton, Ontario, Canada, recognized for his pioneering work in absorptive standing-wave ratio (SWR) bridges, made specifically for QRP operation. The catalyst for using one of his designs came from *An Improved LED SWR Indicator* under VE3DNL's byline appearing in the Spring 2000 edition of the NorCal QRP Club journal *QRPp.*

Of course, the coupler helps your transceiver "see" 50 ohms impedance going into your EFHW, while the SWR bridge allows the operator to "see," via light-emitting diode (LED), when a good match is achieved.

Settling on a Design and Finding Parts

An element from each expert's portfolio makes up the *40- through 17-meter EFHW Antenna Coupler with Super-Sensitive SWR Bridge* design. The schematic is shown in fig. 1. The portion of the circuit beneath SW1 comes from VE3DNL. Everything from T2 to the right is AA5TB's.

There are only a handful of parts needed for this coupler-SWR bridge combination, and many of the parts you may already have on hand. All are available through various parts houses.

Here are some examples:

• **R1-6,** 100 ohm, 1/4-watt resistors (RadioShack® <http://bit.ly/1eSTCRk>)
• **C1-2,** .1-µF capacitors (RadioShack® <http://bit.ly/1bjsyX9>)
• **C3,** 140-pF air-variable capacitor (Hendricks QRP Kits <http://bit.ly/1aCZ1aK>)
• **D1-2,** 1N34A diodes (Dan's Small Parts <http://bit.ly/1bKdKXI>)
• **D3,** LED (RadioShack® <http://bit.ly/1cd49nd>)
• **T1,** FT37-43 toroid (Dan's Small Parts <http://bit.ly/1bKdKXI>)
·**T2,** T50-2 toroid (Dan's Small Parts <http://bit.ly/1bKdKXI>)
• **SW1,** DPDT toggle switch (RadioShack® <http://bit.ly/Iuf3gS>)

For sure, there are many other parts sources for each of the components. By all means, shop around!

Sundry Hardware and Wire

With the exception of nuts and bolts, hardware can be found at RadioShack®:

• Binding posts <http://bit.ly/1d1DlZc>
• SO-239 coax connector <http://bit.ly/1kfFdQp>
• Rubber grommets <http://bit.ly/18FKktA>
• Hook-up wire <http://bit.ly/1dSMs2W>
• Magnet wire <http://bit.ly/1bkdFYV>
• Dual-sided PC board <http://bit.ly/18odZl6>

Once you've built the coupler/SWR bridge, the effort expended in gathering the parts and putting it together will pale in comparison to the enjoyment you'll have in using it.

Let the Soldering Begin!

I decided to construct the unit using Manhattan construction – using small copper-clad pads as solder points for the circuit's components. ***(IN DEPTH:** What is Manhattan construction? Details at* <http://bit.ly/1cXg5em>.*—KI6SN.)*

After scribbling out the point-to-point pad placement based on the schematic, I glued the seven pads on a printed circuit board (PCB) ground plane that had been sized specifically for the enclosure. By the way, the enclosure came from Home Depot. More on that later.

The air variable was mounted at the top of the enclosure where the tuning circuit would be (photo B). The PCB ground plane would go at the bottom of the box—home to the SWR bridge.

When AA5TB's and VE3DNL's circuits were completed, they'd be wired together and connected to the SO-239 coaxial cable transceiver input connector and the antenna (red) and counterpoise (black) binding posts. *(**NOTE:** In fig. 1, input is indicated as XCVR. The antenna is ANT and the counterpoise is CP.—KI6SN.)*

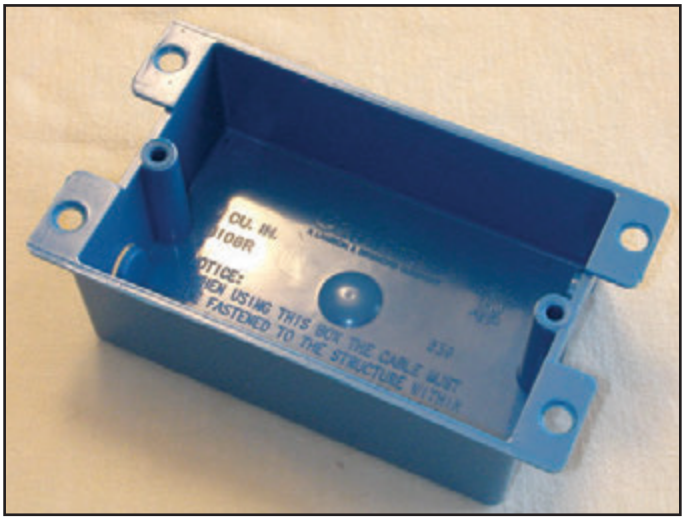

Photo D. Two Carlon 1-Gang, 8 cu. in. Old Work Flanged Shallow Boxes can make a great enclosure for this coupling/SWR bridge unit. Get a pair at Home Depot for less than $3.

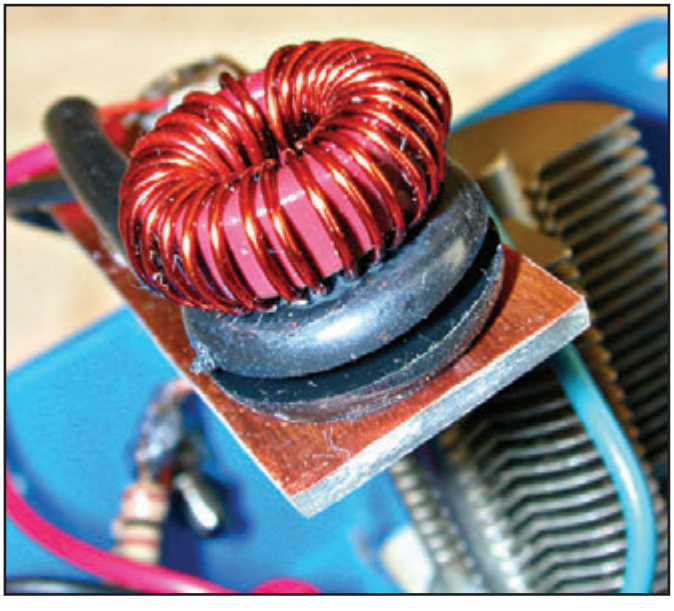

Photo E. Rubber grommets are used not only to insulate the toroid inductors from the PC board ground planes, but to add "shock absorption" to the transformers when the coupler/SWR bridge is being jostled around in a backpack.

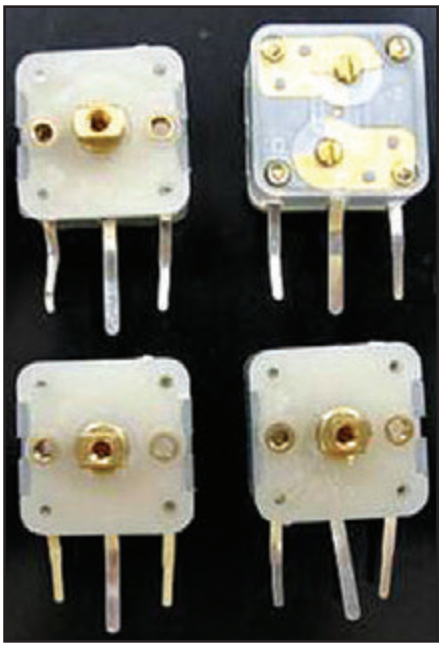

Photo F. There's no need to use an air-variable capacitor for C3. Polyvaricons, each with a 160-pF section, from Hendricks QRP Kits will work beautifully in this circuit. (Internet screen grab)

Since VE3DNL's SWR bridge is *absorptive*, it must be switched IN when tuning for lowest SWR and OUT when tuning is completed. That's the DPDT toggle switch's job. Only then is the operator ready to make contacts.

A view of the completed coupler/SWR bridge is shown in photo C.

What Makes VE3DNL's SWR Bridge 'Super Sensitive?'

VE3DNL branded his accessory *"An Improved LED SWR Indicator."* Here's why:

"To improve SWR sensitivity," he wrote in *QRPp,* "the RF voltage across the bridge arms should be made larger. A broadband RF step-up transformer is required. *(That transformer is made up of the primary and secondary windings on the FT37-43 toroid.—KI6SN.)*

Leinweber pointed out that many LED-based SWR bridge-indicators have a "dead zone" below 2 volts peak-to-peak, resulting in "the sensitivity problem." Back around 2000, "a test with a 2.7-watt transmitter was made to see how high the SWR had to rise before the LED glowed visibly," VE3DNL said. "For SWR below 1.3:1 no light (from the LED) could be seen. With the antenna short-circuited, (infinite SWR), the LED was blindingly bright."

The operator's goal is to tune the circuit so that little-or-no light is coming

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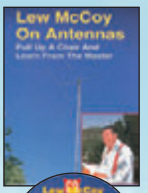
Lew McCoy on Antennas

by Lew McCoy, W1ICP



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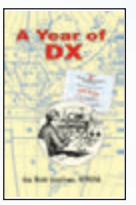

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Photo G. The SWR bridge's LED can be seen recessed behind the grommet to the right of the red ANT (antenna) binding post.

EFHW Antenna and Counterpoise Data

BAND	FREQ.	ANT	CP
40	7.15	65.45	32.73
30	10.125	46.22	23.11
20	14.175	33.02	16.51
17	18.118	25.83	12.92

Frequency in MHz; ANT and CP lengths in feet and decimal feet

Fig. 2. Here are EFHW antenna and counterpoise lengths for 40 through 17 meters. Calculations were based on the frequency at the center of each band. (Courtesy of KI6SN)

from the LED, indicating an impedance match. *Therefore, dead zones such as described are just not helpful.*

"When the 1:10-ratio step-up transformer was added," Leinweber continued, "the same sensitivity test was run. For SWR above 1.06:1, light could be seen." *Much better.*

Leinweber assured that users need not worry about burning out the LED from too much current. In his "improved" circuit, "you actually get less light," he wrote. "With lots of LED current flowing, the RF impedance of the primary winding is very low. It is much lower than the resistive bridge arms."

As a result of this mismatch of impedance, "less RF power is transformed into LED power," he said. "The LED should survive infinite SWR, even when driven by a *full spoonful* (5 watts RF output)."

Project Notes

Enclosure: The piggy-backed plastic enclosures housing the coupler/SWR bridge were $1.20 each at Home Depot and are a mouthful if you ask for them specifically: *Carlon 1-Gang, 8 cu. In. Old Work Flanged Shallow Box.* They're in the electrical department, (See photo D.)

Note that wing nuts are used to hold the top and bottom of the piggybacked enclosure together. This is just personal preference. The two sides are easy to separate if there's ever a need to get inside.

This may sound strange, but: I put several Band-Aids® inside the unit. That way I won't have to think about having a few on hand when I'm on the trail. They don't hurt coupler or SWR bridge performance. And if there's ever a cut or scratch that needs to be attended, I can pop open the box and grab 'em. *Let the wing-nuts fly!*

Resistors: *Why are the 100-ohm R1-6 resistors in three parallel groups?* We want 50 ohms of resistance in each of the three legs. Anyone who has tried to find 50-ohm resistors knows they can be pretty scarce. So by putting two 100-ohm resistors in parallel, we achieve 50 ohms resistance (*remember Ohm's Law—ed.*). Since each resistor is rated at one-quarter watt, that gives us one-half-watt power dissipation in each leg, as well. It's never good practice to "key down" for more than a couple of seconds when tuning up, so the half-watt rating is sufficient at QRP levels.

Inductors: There are two transform-

ers wound on toroids in this design, T1 and T2. Number 26 magnet wire is used for both the primary and secondary windings of each. Note that the 3-turn primary winding of T1 is wound over the middle of the secondary winding. The 3-turn primary of T2 is wound over the *ground end* of the secondary.

When winding is completed, T1 and T2 are glued to rubber grommets, which in turn are glued to the surface of the PC board. *Why?* The grommet separates the inductor from the copper ground plane and acts as a "shock absorber," as well—a nice feature for operators who like to operate trail-friendly radio, photo E. Things can get banged around quite a bit in the field, you know!

Variable capacitor: At KI6SN, a 140-pF air-variable capacitor was used at C3. I'll admit: that's overkill. Hendricks QRP Kits has 160-pF polyvaricon capacitors that are much less-expensive than an air variable. At QRP levels, they will work perfectly well in this circuit. Visit <http://bit.ly/1aCZ1aK>, photo F.

LED: A RadioShack® 5mm Red Wide Angle LED (RS 276-309) was used in the KI6SN version. By all means, though, experiment with any LED you have lying around. It's important to orient the LED properly in the circuit. When holding an LED in your hand, you'll note it has one lead longer than the other. That distinguishes its *cathode* from its *anode.* In fig. 1, the LED's long leg (anode) goes to the junction of D1-C1. The short leg (cathode) goes to the junction of D2-C2.

LED position: You may be wondering why the LED is not pushed through the grommet so it protrudes from the box at the right of the red binding post, photo G. Two reasons:

• If it protrudes, the LED could be damaged when the unit is knocking around on the trail.

• By recessing the LED, it's much easier to see when you're tuning in bright sunlight.

Make it your preference, though—*protrude if you'd like!*

The Easy Part: Making the ANT and CP

OK, it's time to put this jewel in line and on the air. The 50-ohm coax from your transceiver goes to the SO-239 antenna connector on the coupler/SWR bridge.

Now, on what band would you like to operate? Thirty meters might be interesting. Go to fig. 2 and note the figures:

At 10.125 MHz, the middle of the 30-meter band, an EFHW wire must be 46.22-feet long. For simplicity, we'll round that to 46.25 feet, or 46 feet, 3 inches. Pull out your roll of No. 22 insulated stranded hookup wire, measure and cut. *There's your antenna. (**CHECK:** RadioShack® 90-Ft. 22 Gauge Hookup Wire* <http://bit.ly/18p5jOC>.*—KI6SN)*

Referring again to fig. 2, note that a counterpoise for an EFHW 30-meter wire is 23.11 feet—or about 23 feet, 1 1/2 inches. Again, grab that roll of No. 22 stranded wire, measure and cut. *There's your counterpoise.*

Connect the 30-meter 46-plus-foot EFHW wire to the antenna (red) binding post and the corresponding 23-plus-foot counterpoise to the black post. Stretch the counterpoise from the coupler along the ground as straight as you can. Now put the EFHW antenna as straight and high as you can. *(**CAUTION:** Whether you're outside or inside, take care that these wires do not come in contact with any electrical wires, appliances and so on.—KI6SN)*

Whistling a Happy Tune-Up

Once the antenna and counterpoise are in place, it's time to match the antenna to your transceiver.

• Connect your coax from the transceiver to the coupler.

• Flip the DPDT switch to the TUNE position—the side of the switch that incorporates the SWR bridge.

• Find a clear 30-meter frequency and *briefly* key down on CW using the least amount of power possible to get an indication on the LED. You can increase to 5 watts if you like *after* the tuning procedure is completed.

• During key down, chances are the LED will be shining brightly from inside the grommet.

• In no more than a few seconds, turn the variable capacitor (C3) until the LED either dims or goes out completely.

• *Key up!* You've found the matching point.

• Flip the DPDT switch to OPERATE, bypassing the SWR bridge.

Whistle a happy tune! QRO (increase power)—up to a full *QRP spoonful*—if you'd like, because you're ready to get on the air. *Halleluiah!*

Return to fig. 2 and cut another EFHW antenna and counterpoise for another band. And another. And another. *What fun!*

Now It's Your Turn

It's a pretty simple antenna system, wouldn't you say? If a matching unit and SWR indicator are the only things standing in your way, build the *40- through 17-Meter EFHW Antenna Coupler with Super-Sensitive SWR Bridge*. You won't be disappointed, and with such a nice slice of amateur spectrum within reach—either on the trail or in your condominium or apartment building—you may be amazed by your success! I certainly was.

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So here we are at the supposed peak of the weakest sunspot cycle in decades. What's a QRPer to do when the HF bands aren't open? WA6ARA has one suggestion: Look up!

Satellites and the QRPer . . . or Don't Worry About Sunspots

BY MIKE HERR,* WA6ARA

Imagine for the moment a new band, with propagation out to about 3000 miles. Band openings are brief but extremely predictable. Further, it is perfect for QRP operation. You tune across the new "band" at the appointed time and, suddenly, there are signals, some operating in contest style, others in a ragchew. CW and SSB can be heard. You join in, work many stations or just chat with a friend, all with modest antennas and low power levels. And again, at a predictable time, the "band" closes. The "band" I am describing does exist. It's called amateur radio satellites.

Amateur radio satellites, or OSCARs (Orbital Satellite Carrying Amateur Radio), are satellites orbiting above the Earth that contain equipment to relay amateur radio signals great distances. They represent a solution to poor propagation, the need for high power, and the need for large antennas. Many hams consider the use of amateur radio satellites as "fringe," only for the extremely technically inclined amateur, requiring high power, massive antennas, and intricate rigs. This is simply not so. The equipment is readily available, antennas are small, and the required power levels are low. In fact, low power is preferred. Many hams, including QRPers, already have some or all of the equipment that is needed.

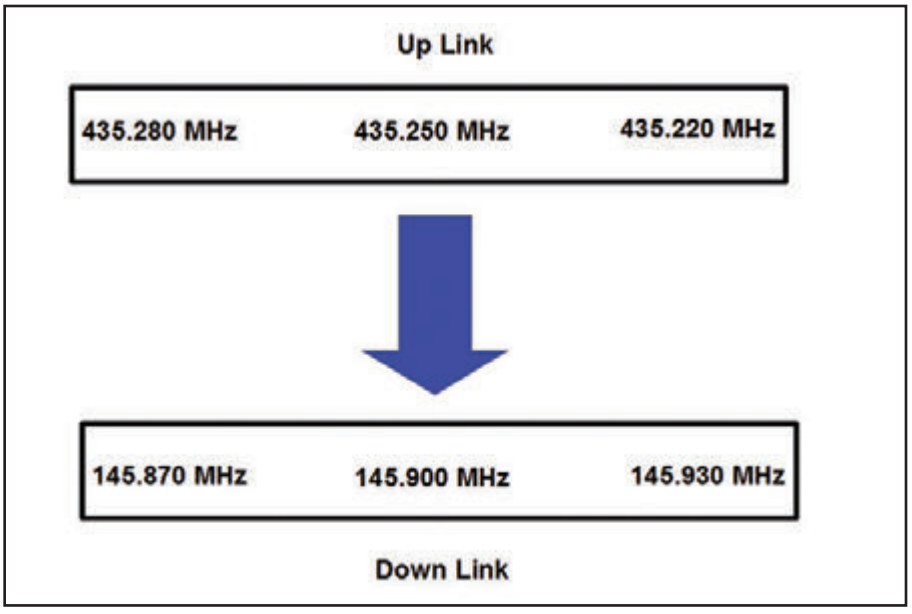

Fig. 1. How a satellite linear transponder works. All signals coming in on the uplink are retransmitted on the downlink. Note that this is an inverting transponder, so the "top" of the uplink passband is the "bottom" of the downlink passband, and what goes in on lower sideband comes out on upper sideband!

Different Satellite "Flavors"

There are basically three types of amateur radio satellites. The first are those doing some sort of scientific mission and using amateur radio frequencies to transmit their data. While it's interesting to track and decode the telemetry, these satellites are not useful for amateur radio communication (although some are opened for amateur use once the scientific mission is completed).

Next are the FM and digital satellites. These are repeaters and digipeaters in the sky. Like their terrestrial cousins, they allow only one user at a time, on a single frequency. These are the so-called "easy sats." You can work them with a handheld, *if* everything is optimum. SO-50[1] is an example of an FM "easy sat."

However, the main interest for the QRPer is the group of satellites containing *linear transponders*. The linear transponder (see fig. 1) takes a portion of one band, amplifies, filters, and heterodynes it to another band. SSB and CW are the primary modes. FM could be used but is *strongly discouraged* due to the bandwidth and 100% duty cycle.

Footprints

Each satellite has a *footprint*, that area of the Earth that it can see. The size of the footprint depends upon the orbital height of the satellite. Anyone inside the

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e-mail: <wa6ara@gmail.com>

footprint may communicate with anyone else inside the footprint. How long you can access the satellite depends upon its orbital speed, which is inverse to its height—i.e., the higher the satellite the slower its ground-path speed and the longer its visibility over any given point on the planet.

High Earth orbit satellites such as AO-10, AO-13, and AO-40 could effectively see half the Earth at a time and would do so for hours. Unfortunately, those magnificent birds are no longer functional. Our present crop of satellites has a more muted footprint and shorter accessibility times. The satellites are constantly moving, so your mutual window with any particular station may only be 5 to 10 minutes long. Fig. 2 shows the footprint of satellite AO-7 on a typical pass.

Amateur radio satellites are QRP-friendly, and low power is encouraged. On the linear transponders, an excessively powerful signal uses a disproportionate amount of the total power available to the transponder. This can cause distortion, drive other signals to become weaker, or even cause the satellite to shut down.

How much power do you need? This depends upon your antenna system. With a modest antenna and a 5-watt signal you can often work across the country. Photo A shows Dave, KB5WIA, having a nice roundtable QSO with W7JPI, VA3JW, WA6DIR, and XE1AO on satellite AO-7. The location is one of the most remote parts of the continental USA, Racetrack Valley, within Death Valley National Park. Note that Dave is using two Yaesu FT-817 transceivers barefoot for the QSO (see below for the reason two radios are needed). Often on a high-angle pass of the satellites I have had great QSOs with only 2 watts of power.

Station Equipment

Many contemporary HF radios have VHF and UHF capability. The FT-817, long a favorite of the QRPer, or the ICOM IC-706, each contain much of the capability you need to operate the satellites. The Elecraft KX-3 with the 2-meter module will also make an excellent portion of the station. If so inclined, as many QRPers are, equipment may be homebrewed.

At a minimum, you will need one radio to transmit and a second one to receive, unless your radio can simultaneously receive on one band and transmit on another. It is very important to be able to listen to your signal. Why? First, to make sure you are not using too much

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Photo A. Dave, KB5WIA, having a roundtable QSO on the AO-7 satellite from Death Valley National Park in California, one of the most remote parts of the continental USA.

power! You don't want to be the "lid" who causes the satellite to distort or shut down. QRPers will not have this problem! The second reason is to track your own signal. The satellite is moving, a lot, on the order of 17,000 mph. While operating, the signal will seem to drift. It isn't; it is the Doppler Effect.

As the satellite approaches your station, the frequency will be high, and as it passes and recedes, the frequency will drop. The same can be observed with a passing train. The sound of the horn changes frequency as it passes. By being able to separately adjust the receive and transmit frequencies, and hear your own signal, you can compensate as well as knowing if you are being heard.

If you have a computer and CAT-controlled radios, then the computer can adjust the radios for the Doppler shift. Otherwise, for manual control, the rule

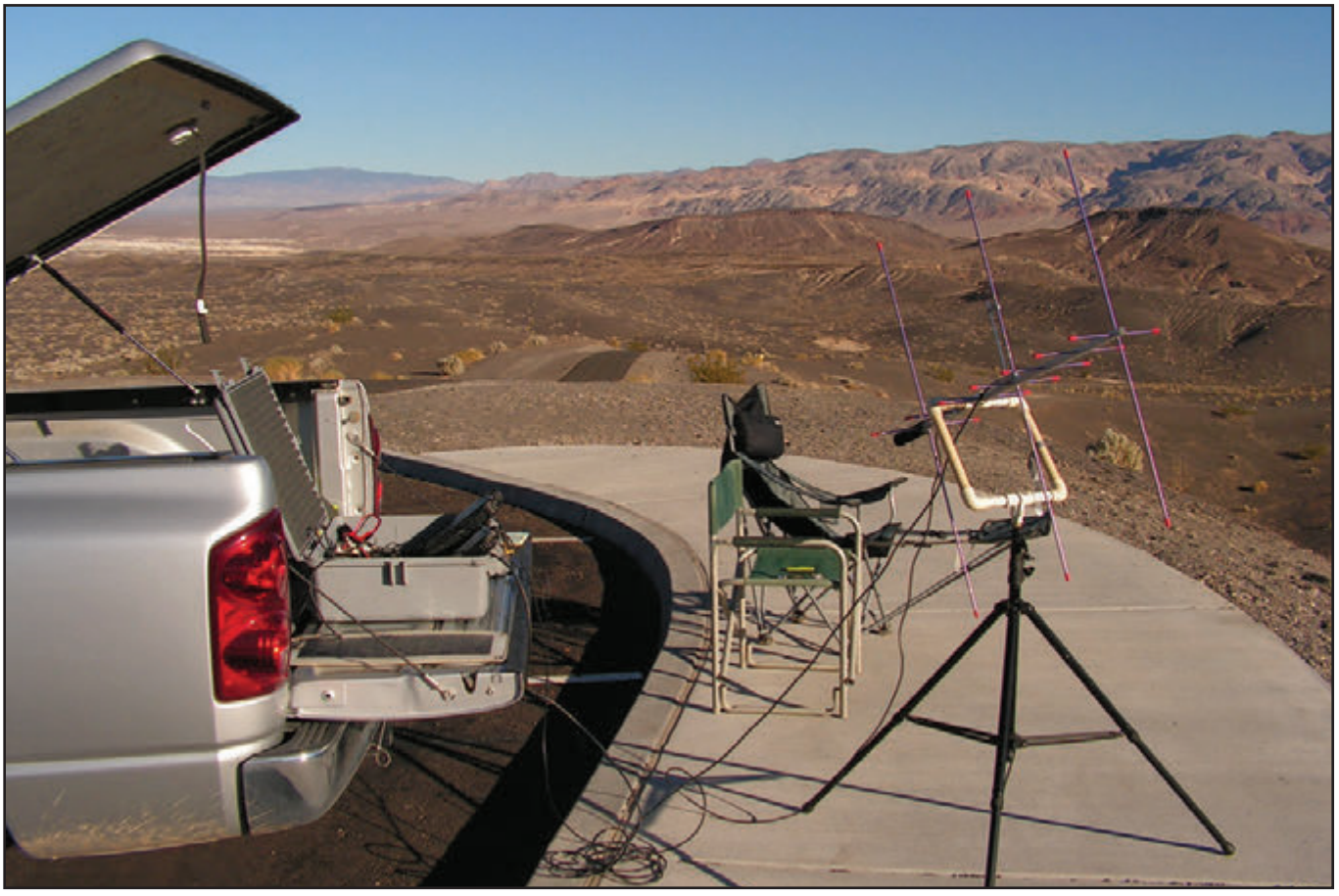

Photo B. The author's portable satellite setup at the edge of Ubehebe Crater in Death Valley, California.

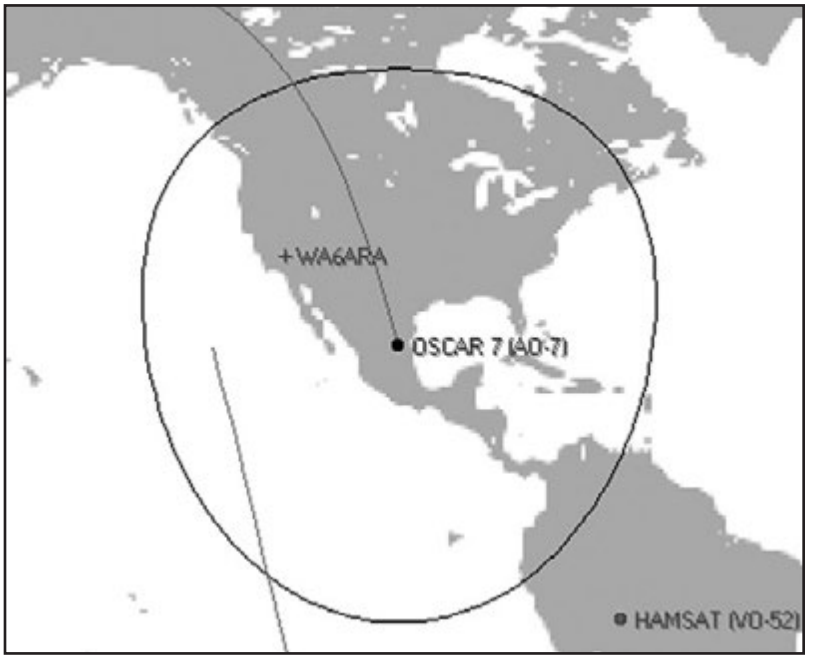

Fig. 2. "Footprint" of OSCAR-7, a nearly 40-year-old satellite that is still going strong. Any two stations within that "footprint" may communicate with each other via the satellite.

of thumb is to adjust only the *higher* frequency of the uplink or downlink. For example, on VO-52 you transmit on 435 MHz and receive on 145 MHz. Leave the receive frequency alone and only adjust the transmit frequency. Mark Spencer, WA8SME, has developed a way to computer control a single FT-817 for both transmit and receive, and to compensate for Doppler.[2] While you will not be able to hear yourself, at the 5-watt level you are not at risk of exces-

Where Are the Satellites?

There are a couple of ways to track amateur radio satellites, either online or with a computing device. Just about any computer will work … desktop, laptop, tablet or phone. Various programs and apps exist, some free, some for a nominal fee. All work well. Also, you will need what is called the *Keplerian elements* for the particular satellite of interest. These are the mathematical values which describe the orbit of the satellite, originally developed by Johannes Kepler in the 17th century. (Of course, he had no idea at the time that his calculations would someday be applied to manmade satellites.)

Online tracking has the advantage of not having to worry about the latest Keplerian elements. AMSAT has an excellent quick look tool at <www.amsat.org/amsat-new/tools/predict/>. Another one is at <www. heavens-above.com>. It has a specific tab for amateur radio satellites. Just open it, enter your location and select the satellite you want. You can get the pass info for several days if desired. The downside is that it only gives you the start location of the pass, the highest point and then the end-of-pass location. It is up to you to mentally interpolate the times and points in the sky as to where the satellite is during a pass. It sounds hard, but really isn't if you are moving the antenna manually. Another negative to online tracking is that you can't slave a rotor or radios to automatically follow the satellite in terms of location or Doppler shift.

If you want to use a computer, then there are a number of options available. For Windows®, the program **Satpc32** is considered the best and has a large support base. It is available for a nominal fee at the AMSAT store <http://store.amsat.org/catalog/index.php?cPath=2>. Also at the AMSAT store is **MacDoppler** for Mac users.

In the freeware world is the excellent program **Orbitron** at <www.stoff.pl>. Linux users should check out **gpredict** at <http://gpredict.oz9aec.net/>. All of these allow for the automatic updating of Keplerian elements as well as slaving the rotors and radios to the computer for automatic tracking and Doppler correction.

Tablet and phone users, especially those with the Android system, should check out **AmsatDroid** at the apps store. While you can't drive the rotors or radios with it, it does give you a nice listing of available passes as well as a world map showing the satellite's location. For the iPhone user, try **HamSat** or **ProSat**, available at the iTunes Apps store.

Your First Satellite Contact

Your first satellite contact, just like your first HF or VHF contact, can be very trying, even for a seasoned ham. I would recommend reading the "Getting Started" series at the AMSAT website <http://www.amsat. org/?page_id=1869>. While somewhat dated, it is still good information.

If the FM satellites are what you are interested in, then a good tutorial for SO-50, presently the only FM satellite we have, is at <www.g6lvb.com/articles/operatingso50.htm>.

For the linear transponder "birds," here is a plan:

First, I will assume you have the antennas and the radios ready to go. I am also going to assume that for your first contacts, you will be doing everything manually. What satellite to go after? I strongly suggest VO-52. It is very strong and reliable. Look at the available passes for your location. Select a time and pass with a high degree of success probability. Weekends tend to be more popular than weekdays; evenings tend to do better on the weekdays and daytime on the weekends. Think about who your audience might be. If you are on the west coast, don't select a pass that is largely over the Pacific Ocean. And don't pick a low-angle pass. While these hold the promise for the most DX, a nice high-angle pass, say above 45 degrees, will be easier for getting started.

Next, get everything ready at least 15 minutes prior to the start of the pass. You don't want to be searching for a coax adapter as the satellite comes up over the horizon. Know which way north is. That sounds simple but you want to know exactly which way *true*, not magnetic, north is as well as other compass points.

For this first pass, I suggest you only listen, do not transmit. Why? All of this is new—tracking time, azimuth and elevation angles as well as frequency—so eliminating transmitting will help out the first time. Select an open area, preferably one without trees or buildings to block your sky view. Have the pass information in front of you, receive radio all powered up, set to upper sideband (more on this later), and the antenna pointed at the approximate location of the beginning of the pass.

The appointed time has come! Start tuning around the satellite's downlink passband, mostly at the center. For VO-52 this is 145.900 MHz. Tune up and down about 15 kHz. Soon you will hear a voice or two, or maybe a CW signal. Move the antenna in both azimuth and elevation to maximize the signal, but keep your eye on approximately where it should be in the sky from the tracking information. More than once I have found myself tracking a satellite using a minor lobe of my antenna and wondering why it is so weak.

Once you find a QSO, listen and see how it is done. Typically, you will find two different types: the fast contest-style QSO and the rag chew. Sometimes the rag chew turns into a big round-table with stations checking in as the satellite comes within their view and others falling off as it leaves. You may have to touch up the receive frequency a bit during the pass. When the pass is over, it is over.

Once you have mastered the receiving end of things, it's time to transmit and hold a QSO. The setup is the same except for the transmitter. For VO-52, you will set it up at 435.250 MHz, lower side band. Why LSB? Because the transponder in VO-52 is *inverting,* so what goes in as lower comes out as upper, and upper sideband is the convention for satellite downlinks. The rest of the setup is the same as receiving.

Now the pass begins, and you hear someone calling CQ. The hard part is where to transmit. First, because the satellite inverts the signal, as you tune *up* in frequency on the uplink (435 MHz) it goes *down* in frequency on the downlink (145 MHz). And just to confuse you a bit more there is old Doppler messing with you. So I suggest you first go down to say 145.890 MHz and tune to 435.260 MHz transmit and send your call sign in CW. Tune with the receiver until you hear it. Once found, only tune the transmit side, not the receive side, of your radio set up. You might try this for a pass or two until you get the hang of finding where you are and correcting for Doppler. In fact, a good time to do this practice is at the times when you would not expect much traffic on the satellites, i.e. over the oceans or mid-day during the week.

Once you have the hang of it, try answering that CQ. Or call CQ yourself. Warning, it may take a time or two to get everything down with moving the antennas, correcting for Doppler and using two radios, but it will quickly become second nature. Adding computer control for antenna positioning and Doppler correction greatly reduces the overall workload and can make satellite operation a real pleasure. Enjoy!

The Next Step. After you have your feet wet on satellite operation, *become the DX*. Satellite operators love to chase Maidenhead grid squares. And the rarer, the better. I am in DM15, fairly rare, but I am on often, so it's not a big problem. I can drive a couple hours to Death Valley, though, and be in DM16 or DM17, which are rare indeed. When you get on the satellite from a rare grid, you can expect a pile up, as everyone wants your grid! Many of us live within a couple hours of grid squares that are rarely, if ever, activated on the satellites. So a day trip with a portable satellite station can be a lot of fun!

sive power and the Doppler correction is done for you. It makes for a nice single-radio package.

Satellite Antennas

Antennas really are the secret to successful satellite operation, just as they are for HF QRP work. For satellites, this means beams. Fortunately, the beam antennas necessary for 145- and 435-MHz operation (the primary bands for satellite operation) are relatively small and can be near ground level. Many operators use a simple tripod from the back yard or balcony, so there's no obvious antenna for the homeowners' association to get excited over.

For the homebrewer, *CQ* Antennas Editor Kent Britain, WA5VJB, has an excellent design for a 145-MHz/435-MHz antenna for satellite use.[3] It is effective, cheap, and simple to build. Commercial antennas by Elk[4] and Arrow[5] have been successfully used for portable and permanent QRP satellite operation. The Elk antenna is a 5-element dual-band log periodic. The Arrow antenna consists of a 3-element 145-MHz antenna with a 7-element 435-MHz antenna set at cross polarization on the same boom. I use one of these in the field with excellent results. Photo B shows my simple setup using an Arrow antenna at the edge of Ubehebe Crater, in Death Valley, California.

It is necessary to know when a particular satellite will appear and what its "track" will be as it crosses the sky. This can be worked out with home computers, tablets, or smart phones. Several excellent programs are available, both commercial and freeware. Once the satellite's pass is known, then the antennas need to track or follow the satellite during its "pass" over your location. At the home QTH, this is often done with rotors in both the azimuth (compass direction) and elevation (up and down) axes. The rotors may be driven by the computer, freeing up the operator of the tracking task. In the field, it is most often done by hand. Simply point the antenna at the approximate location in the sky and tweak slightly for best signal. A helper here is best, as the operator can easily become overtasked with tracking, frequency correction and operating.

Linear Transponder Satellites

There are presently three satellites in orbit with linear transponder capability. These include AO-7, FO-29, and VO-52. AO-7 is the granddaddy of them all. Launched in late 1974, it died in 1981, then amazingly came back to life in

2002. It operates only in daylight (when its solar panels are gathering energy) due to open batteries. The fact that a nearly 40-year-old satellite is still in service speaks highly of solid amateur radio engineering.

AO-7 contains two sets of transponders, one for 145 MHz up and 29 MHz down, and another for 432 MHz up and 145 MHz down. It cycles back and forth between the two transponders approximately every 24 hours. AO-7 is more susceptible to high-power distortion due to the lack of functioning batteries and QRP is very much welcomed here.[6]

FO-29 is another workhorse satellite.[7] Launched by JAMSAT (Japanese AMSAT) in 1990, it still performs well. FO-29's uplink is on 145 MHz and its downlink is on 435 MHz.

VO-52 was launched by India's AMSAT organization in 2005. This is a very strong signal satellite.[8] It receives on 435 MHz (uplink) and transmits on 145 MHz (downlink).

Looking in the Crystal Ball

What does the future hold for amateur radio satellites? Are you going to build a station just to have it be idle due to lack of satellites? No way! At the time of writing this article, several satellites which include linear transponders had just been launched. Their transponders will be activated once their primary scientific missions are completed, maybe about the time you are reading this. These include FUNcube-1 and Delfi-n2Xt.

This year should see the launch of FUNCube-2, KiwiSat, CAS-2A1, and CAS-2A2. In addition, a plethora of satellites with FM repeater capability are scheduled to be launched. AMSAT DL in Germany is actively working on an OSCAR-13 class satellite and there is word of AMSAT NA initiating work on a small "cubesat" satellite with high Earth orbit potential.

So the future is indeed bright. Check out the AMSAT NA web page (www.amsat.org) for specific information on these and future satellites, and consider joining AMSAT and supporting amateur radio satellites.

I have only touched on the use of satellites and QRP. Using satellites, the QRPer opens up a whole new field of operating. Power levels need not be high, antennas may be small and out of the way, homebrewing is encouraged, and the operating in the field, which many QRPers also enjoy, is perfect for including satellites. By all means, get on and give them a try!

Notes

1. See <www.g6lvb.com/articles/operatingso50.htm>
2 - Spencer, Mark, WA8SME, "Satellite CAT Interface for Working the Analog Birds," *QST,* Oct 2012
3. <http://www.wa5vjb.com/references/Cheap%20Antennas-LEOs.pdf>
4. <http://elkantennas.com>
5. <http://www.arrowantennas.com>
6. See <http://www.amsat.org/amsat-new/satellites/sat_summary/ao7.php> and <http://www.satview. org/?sat_id=07530U>. *(Note: While the AMSAT page cautions users about questionable legality of transmitting to AO-7 due to rule changes in the past four decades, the satellite is in regular use and we are unaware of any enforcement action ever being taken against any user.—ed.)*
7. <http://www.amsat.org/amsat-new/satellites/sat_summary/fo29.php>
8. <http://www.amsat.org/amsat-new/satellites/sat_summary/hamsat.php>

BY IRWIN MATH,* WA2NDM

How to Properly Listen to a Restored Antique Radio

Last month, as you may remember, we described the restoration of an antique Atwater Kent radio receiver we obtained at an extremely low price at a local flea market. Well, now that we had a working 1930s era receiver, the thought that if we could listen to some of those actual radio shows of yesteryear on the radio would truly be nostalgic. To do this, however, we needed two things—a recording of some old shows and a suitable transmitter to send them to the radio. As it turned out, the old-show recordings were readily available on the internet (at very reasonable prices) and we purchased a CD with several of our favorites. The transmitter was another matter.

This month we will describe what we came up with. Before starting, you should also realize that this same technique can be used to build a simple low-power (QRP) AM transmitter for almost any amateur band.

For the most part, old radios were AM receivers covering the broadcast band from around 550 to 1500 kHz (1.5 MHz). What we needed, therefore, was a small, low-cost AM transmitter with a range of 10 to 20 feet. Our plan was to have the transmitter hidden in another room, modulated with the old radio show recordings and then fool our guests by telling them that somehow our old radio could pick up stations from the past! To keep things legal we had to limit the power of the transmitter to less than 100 mw into a 10-foot antenna to properly comply with FCC regulations. In the past this would have been done by means of a so-called vacuum-tube-operated "phono oscillator," but we wanted something much more modern and battery operated if possible.

Our first attempt was to use a surplus 1-MHz oscillator (ECS 2100AX-1.0MHz available from Mouser Electronics for less than $5) that we had on hand. This was a TTL unit normally intended for microprocessor clock applications, but since its output was pretty much in the middle of the broadcast band, it seemed like an ideal choice. Applying 5 volts, a 1-foot length of hookup wire as an antenna a foot or so from the radio, and tuning the radio to "100" on the dial resulted in a strong carrier. Now to try to AM modulate it.

Since the output of the oscillator would vary in amplitude from about 3 volts to 5 volts depending on the input DC provided, our first attempt was to simply vary the supply voltage to the oscillator by means of a 1K to 8-ohm audio output transformer that we had on hand. The transformer, a RadioShack 273-1380, was connected backward in the classical plate modulated circuit shown in fig. 1. We then applied a 400-Hz tone from our signal generator to the 8-ohm winding, our scope to the output of the oscillator, and then adjusted the input level until we got a sine-wave modulation envelope. The first waveshape shown in fig. 2 was what we could get, but the second waveshape is what was really desired. When actual audio was applied to the transformer, the sound was clearly present when received by the radio but somewhat distorted, probably due to the non-symmetrical downward modulation. In addition, there were several harmonics due to the square-wave output. Not perfect, but usable.

To clean things up we added a class C tuned amplifier stage to the output of the oscillator as shown in fig. 3. We also moved the modulation transformer from the oscillator to the amplifier, thereby achieving the final modulation waveshape we desired. This stage not only increased the output amplitude but also cleaned up the wave form to make it more sinusoidal and reduce higher order harmonics. In addition, you will note that we con-

*c/o CQ magazine

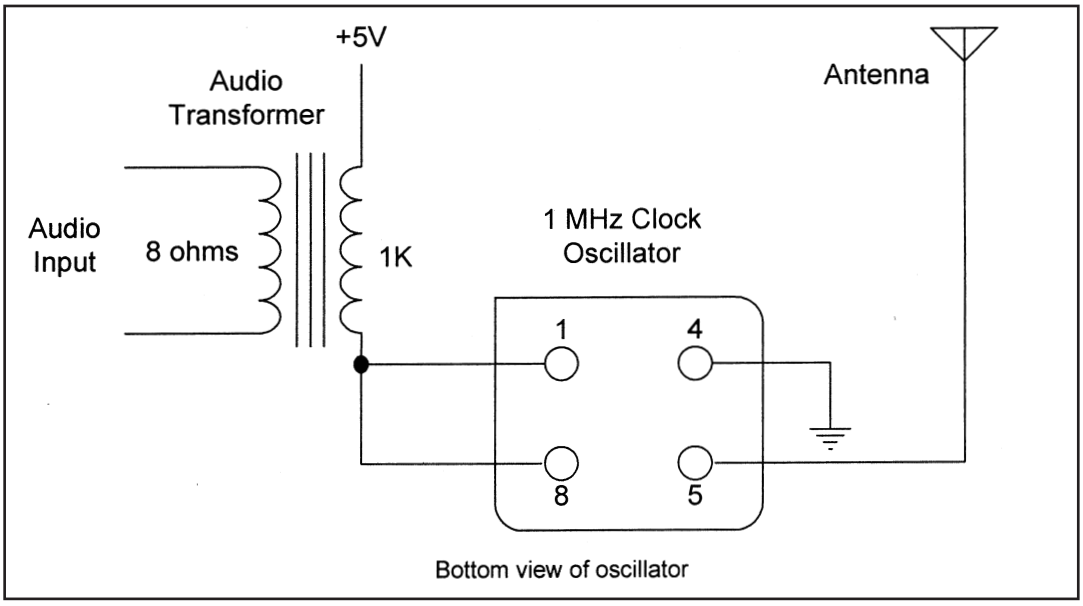

Fig. 1. Basic modulated AM "transmitter."

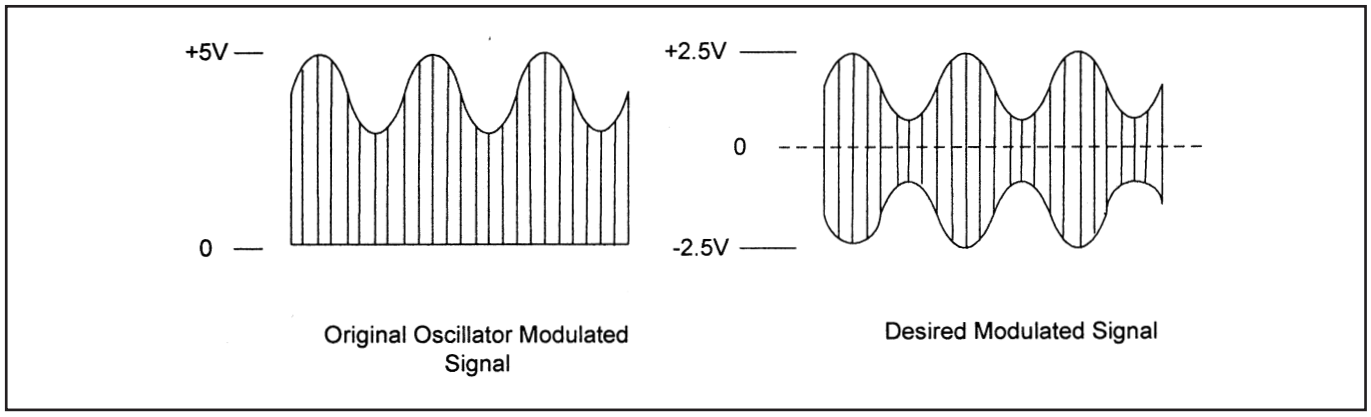

Fig. 2. Waveshapes from "transmitter."

nected the amplifier stage to 9 volts (for more output) and added a 3.9-volt Zener diode to drop the 9-volt supply to the 5 volts required by the oscillator. Note Ct and Lt. These values were chosen to resonate at approximately 1 MHz and were what we had on hand. You certainly can experiment with these values to peak the output signal as much as possible. When you do, adjust the values (and audio input level) as required to achieve a symmetrical modulated signal with no clipping (see fig. 2). Also, whenever using this type of circuit be sure to use only as much antenna length as you need to go a short distance. Too much and your audience may be the FCC!

As we mentioned before, you can use this same technique to build a low-power AM transmitter for any amateur band where you can purchase an oscillator with the correct frequency. Simply scan the catalogs of the various electronic suppliers and look for the frequencies available for clock oscillators. Then adjust the values of the tuned circuit as required for best results. You may be able to change the output transistor for a higher power device to achieve more output, and "key" pin 1 of many similar oscillators for CW operation, but that is a topic for another time. As a final reminder, be sure to check the data sheet for the oscillator you use. You will find additional information as well as any changes in wiring required for the particular unit you may have.

Our final result was a real nostalgic experience. What helped even more was that the old-time radio program CD we obtained also had commercials from the same time period. Under the right circumstances (if you were not in the 21st century) you would swear you were back in the golden age of radio! 73, Irwin, WA2NDM

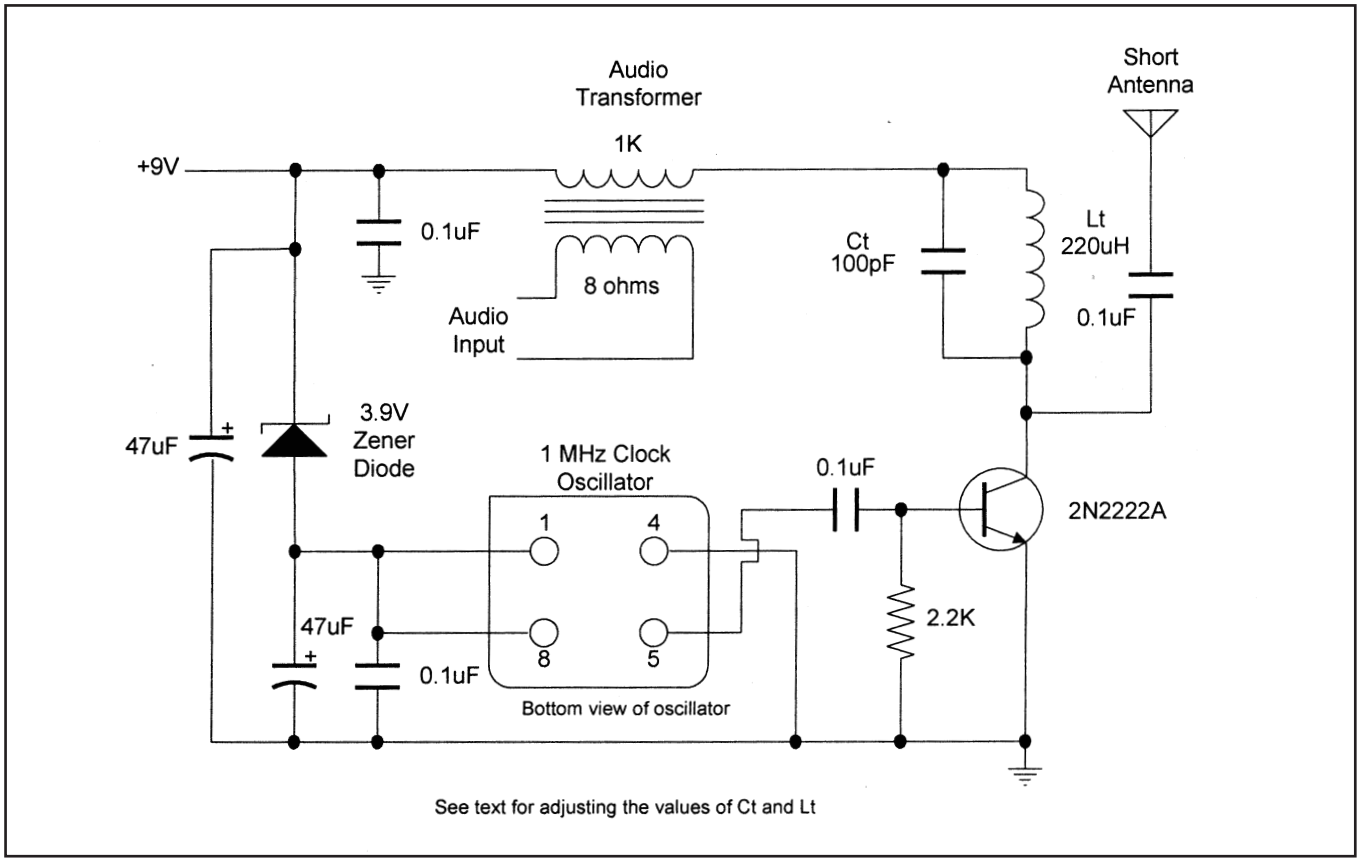

Fig. 3. Final modulated AM "transmitter."

BY RICHARD FISHER,* KI6SN

EmComm in the Sky and on the Ground

SKYWARN Hails Volunteers, while "Spokespersons" get Support from Florida Hams

While a large portion of the Midwestern and eastern United States were heading into a freak pre-winter storm December 7–8, it was "82 degrees and sunny" at the National Hurricane Center in Miami.

If you're an NFL fan, you remember the weekend "Snow Bowls" from Philadelphia to Washington, D.C.; Baltimore to Pittsburgh.

owever, the balmy WX (weather) report from South Florida was what operators at the NHC's WX4NHC issued in QSO exchanges during the annual SKYWARN® Recognition Day (SRD), an all-Saturday on-air activity in commemoration of the organization's thousands of volunteers (photos A and B).

"This is a nationwide on-the-air communications exercise that recognizes volunteer SKYWARN spotters for their dedication and skills in providing the National Weather Service (NWS) and their communities with real-time surface reports during severe weather," Julio Ripoll, WD4R, said. For more than three decades he has been WX4NHC Amateur Radio Assistant Coordinator. "These surface reports help NWS forecasters and local authorities provide the public with advance warnings of severe weather."

SRD is sponsored by Scott Mentzer, NØQE, and Matthew Mehle, KCØTER, of the National Oceanic and Atmospheric Administration (NOAA) <http://www.NHC.NOAA.gov >.

According to the SKYWARN website, it is "a volunteer program with nearly 290,000 trained severe-weather spotters. These volunteers help keep their local communities safe by providing timely and accurate reports of severe weather to the National Weather Service." ***(IN DEPTH:** For complete SKYWARN information, visit <http://www.SKYWARN.org>. – KI6SN.)*

Although spotters "provide essential information for all types of weather hazards," the site notes, "the main responsibility of a SKYWARN® spotter is to identify and describe severe local storms. In the average year, 10,000 severe thunderstorms, 5,000 floods, and more than 1,000 tornadoes occur across the United States. These events threatened lives and property."

During 2013 SRD, radio amateurs from around the world were invited to check in with WX4NHC, giving a brief weather report. It's an opportunity, as well, to test the reliability of each operator's communications equipment.

During the daylong event, "We tested all of our equipment and antennas, including HF, VHF, UHF, EchoLink/IRLP, APRS, and other digital and on-line modes," Ripoll said from WX4NHC.

"Although the HF (high-frequency) propagation from South Florida was not very good, we were able to make 124 contacts on 15, 20, and 40 meters, including 22 NWS offices. The VoIP NWS/NHC Net—run by the SKYWARN group WX1BOX at NWS's Taunton, Massachusetts site—was a success. This net had scheduled many NWS offices and linked EchoLink and IRLP repeaters on Conference room WX-Talk."

Ripoll noted that WX4NHC's most distant contact and weather report was from Tony Langdon, VK3JED, of Long Gully, Victoria, Australia, who was operating mobile using an IRLP repeater.

The lowest temperature reported was from the Arrowhead SKYWARN Response Team, KØNWS, in Duluth, Minnesota. *How cold was it, Johnny?*

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Photo A. Lloyd Kurtzman, W4KAM, left, and Julio Ripoll, WD4R, take their turn at the operating position at WX4NHC from the National Hurricane Center, Miami, for 2013 SKYWARN® Recognition Day. (Courtesy of WD4R)

How about –6 degrees Fahrenheit, with a wind chill factor of –26 . . . *and snowing?*

"We felt a little guilty when we gave them our report of *82 degrees and sunny* in South Florida," Ripoll said.

WX4NHC made contact with WFØEMA, at Federal Emergency Management Agency (FEMA) Headquarters, on 15 and 20 meters with Ted Okada, K4HNL, at the controls from FEMA.

"Ted had visited WX4NHC," Ripoll said, "and then installed the amateur radio station at FEMA Headquarters. This was the first HF contact between FEMA HQ and the NHC."

WX4NHC "appreciates all of the SKYWARN Spotters for their dedication and valuable surface reports," Ripoll said. "They are the eyes and ears for NWS and NHC."

He also thanked the VoIP Hurricane Net team operating from NWS Taunton "for sponsoring the net on EchoLink/IRLP."

Stations that contacted WX4NHC can request a QSL card via WD4R <http://www. QRZ.com/db/WD4R>. *(**VISIT:** The WD4R home page at <http://www.wd4r.com>.—KI6SN)*

Photo B. The Miami-based WX4NHC SRD team included, from left, Julio Ripoll WD4R; Alan Wolfe, WB4L; Phil Moldovan, N4LCZ; and Louis Cruz, N4LDG. (Courtesy of WD4R)

Photo C. Robert McNeill, W4MBD, and Ed Pitts, K5OF, take their post in the Newport, North Carolina NWS conference room during SRD 2013. (Courtesy of KJ4JPE and John Cole)

Grassroots SKYWARN® Activity was Brisk in North Carolina

While WX4NHC is widely regarded as the hub of the wheel during events such as SKYWARN® Recognition Day, there is a beehive of activity at the grassroots level, as well. Here is a snapshot of SRD action centered on the National Weather Service office in Newport, North Carolina.— *KI6SN.*

SKYWARN® Recognition Day 2013 began at 2400 UTC at NWS Newport with the arrival of Ron Civils, KC4WVV, of Deep Run, North Carolina, who spent the wee-hours of Saturday making contacts on 40 meters.

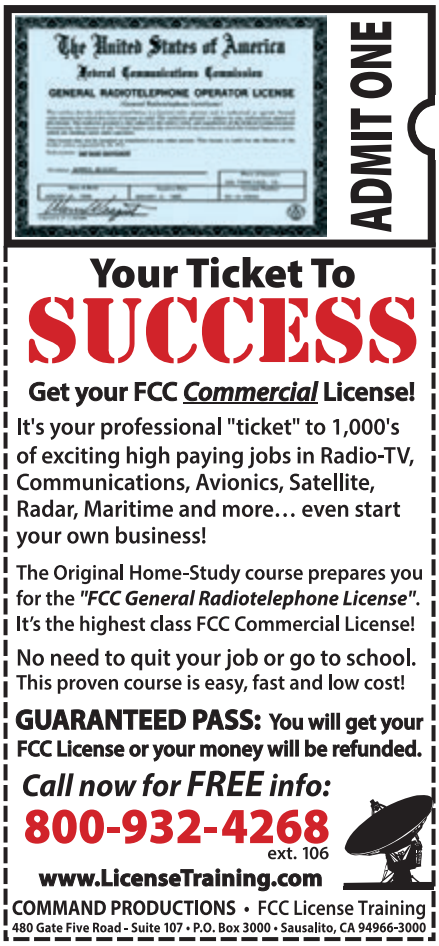

Photo D. William Moore, KK4SUF; David Kay, KJ4IGZ; Terry Passey, K4TRP; and Mike Miller, KK4VNH; oversaw operations from the Carteret County EmComm trailer during 2013 SRD. (Courtesy of KJ4JPE and John Cole)

Meanwhile, from her home in Swansboro, Janice Hopkins, KJ4JPE, made contacts as WX4MHX on the 2-meter K4GRW repeater using IRLP 9219. Later Saturday she carried on that activity from the NWS office in Newport. *(**MORE:** For a peek inside WX4MHX, visit <http://www.QRZ.com/db/WX4MHX>.—KI6SN)*

Terry Passey, K4TRP, made 10-meter contacts throughout Saturday using the Carteret County Amateur Radio EmComm trailer. Working inside NWS Newport was Ed Pitts, K5OF, using his Yaesu HF transceiver on 20 meters with the help of David Myers, KK4DEX, and Ed Myers, K4PEF's portable antenna trailer.

NWS Newport Chief Coordinating Meteorologist John Cole and Meteorologist Hal Austin, NC4HA, joined other staff members in welcoming more than thirty radio amateurs and their families, and giving tours of the facility.

Nineteen NWS contacts were made, along with many other SKYWARN amateurs (photos C and D).

*(**NOTE:** Information for this report was provided by Janice Hopkins, KJ4JPE, ARRL North Carolina Section Public Information Officer. She shared photography duties with NWS Newport's John Cole.—KI6SN)*

Photo E. Richard Pelchat, KK4KTX, maintained radio communications with Lake County ARES® net control station N4FLA from his Mount Dora Bicycle Festival (MDBF) rest stop radio post. (Courtesy of K1AYZ)

Florida Hams "Spoke Up" for Mount Dora Bicycle Festival

The Lake County (Florida) Amateur Radio Emergency Service provided radio communications to support the three-day Mount Dora Bicycle Festival sponsored by the Mount Dora Area Chamber of Commerce.

This was the 39th year for the annual event held each October.

"We had sunny weather for the three days with moderate temperatures and reasonable humidity so there were no reports of heat exhaustion," writes Ted Luebbers, K1AYZ, Lake County ARES® (Amateur Radio Emergency Service) Public Information Officer. "There were 703 bicycle riders registered this year, which was a slight increase over the past year" he said. "Most of the riders were from Florida, but there were some from various places in the United States. This is not a competitive race but a series of 15 separate bicycle tours through Lake County's countryside. The county has 1,400 named lakes and is considered to be Central Florida's hill country. This combination provides a scenic background and significant challenges for the riders. It keeps them coming back year after year."

The ARES group had 20 radio operators helping to keep track of the bicycle riders, K1AYZ reported. ARES also set up at rest areas, provided SAG wagons (vehicles that support cyclists) and radio-equipped mobile patrol vehicles (photos E and F).

"The ARES volunteers provided 255 man hours of service over the three days, Luebbers said. "All vehicles were directed by the N4FLA net control station, which had its mobile command trailer set up in the parking lot of the Mount Dora Chamber of Commerce.

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Best wishes for the holidays and throughout the New Year.

"We used a 2-meter repeater provided by K4AUS on 147.390 MHz, and APRS (Automatic Packet Reporting System), with the 27-inch video monitor mounted in the trailer to keep tabs on the three SAG wagons. This video system can also be used to monitor weather conditions during any event."

Photo F. Strait Hollis, KT4YA, ARES® event coordinator for the MDBF, used his pick up as one of the mobile patrol vehicles and cyclist support wagons during the three-day festival. (Courtesy of K1AYZ)

A total of nine riders were transported to the Mount Dora staging area due to mechanical problems "or just getting tired out," K1AYZ said. "We had one report of *a rider down* with what appeared to be serious injuries . . . EMS was called and the fallen rider was transported by air to a local hospital for evaluation. Later we received a report that he was not seriously hurt and was released from the hospital."

Luebbers noted that the Lake Area Radio Association has been providing on-course radio communications for the Mount Dora Bicycle Festival for more than twenty years. LARA uses events such as this "to train its members to be ready to provide emergency radio communications for disaster situations such as hurricanes and tornadoes, which are not uncommon in this part of Florida. Members volunteer their time, radio equipment, and vehicles as a public service to the community," K1AYZ said.

*(**IN DEPTH:** For more information about amateur radio activity in Lake County, visit <http://www.n4fla.org> and <http://www.k4fc.org>.—KI6SN)*

Finis . . .

That's it for this month's *CQ* "Public Service" column. Remember, we're always on the lookout for the stories of your EmComm activities—whether it's from the scene of an emergency, training for one or your public relations efforts in supporting community activities. Please keep me on your mailing list via <KI6SN@cq-amateur-radio.com>. Every report helps us all to hone our EmComm skills. Let's hear from you or your organization soon!

73, Richard, KI6SN

kit-building

BY JOE EISENBERG,* KØNEB

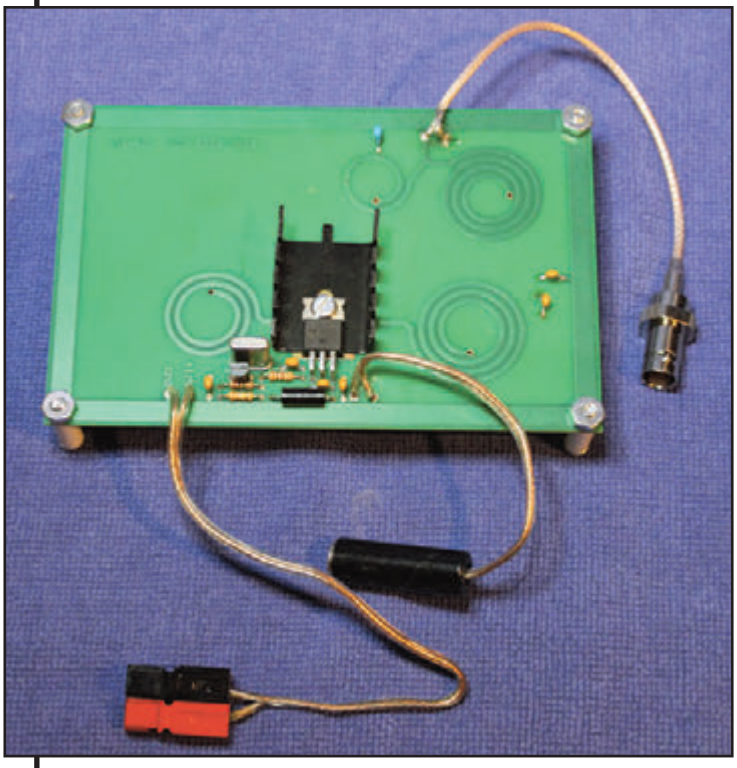
QRP Special

A Pair of QRP Quickies

When choosing a kit for building as a group or a first-time kit builder, the choices sometimes can be difficult. When building a kit as a group, the main limitations of time and space come into play. The need for a simple and quick-to-build kit becomes most important. Here is a kit that meets those requirements and puts out a great CW signal! The NS-40, designed by David Cripe, NMØS and sold by the Four States QRP Group is a great little 40-meter, crystal-controlled, 5-watt CW transmitter.

The NS-40 has a very small parts count, and no toroids to wind due to its unique design, making use of PC-board printed coils integrated into the PC-board design. Of course, the tradeoff with this design is that the board has to be a bit larger than would have been necessary if toroids had been used, but with no toroids to wind, you get a QSL-card-size board (more on that later). With just 14 parts to be soldered onto the board, this kit can be assembled by experienced builders in less than 30 minutes! This makes the NS-40 ideal for a beginning kit builder or for a group kit-building experience. I placed the resistors and capacitors first, followed by the other components.

*7133 Yosemite Drive, Lincoln, NE 68507
e-mail: <k0neb@cq-amateur-radio.com>



Completed NS-40 Kit. Only 14 parts make up this fun CW transmitter!

Unlike most kits, there aren't very many parts to sort, so the main cautions that apply are to be very careful in identifying the various capacitors and be sure that they are mounted in their correct locations. Watch your transistors to be sure they are mounted correctly as well, and be sure to use a little heat-sink grease on the final output transistor when mounting it to the board and its heatsink. The heatsink grease allows a more even and efficient flow of heat away from the transistor to the heatsink. The two transistors are the only polarized components in the kit, not including the wiring.

The PC board for the NS-40 is silkscreened and plated through and pre-tinned, making for quick assembly. What makes this PC board special is not only the built-in coils, but also the bottom of the board is printed in the form of a QSL card. Just fill in the particulars of your first contact, take a photo of it, and send that photo to your first contact! What a unique QSL card, and a great reminder of your very first QSO using this simple and fun kit.

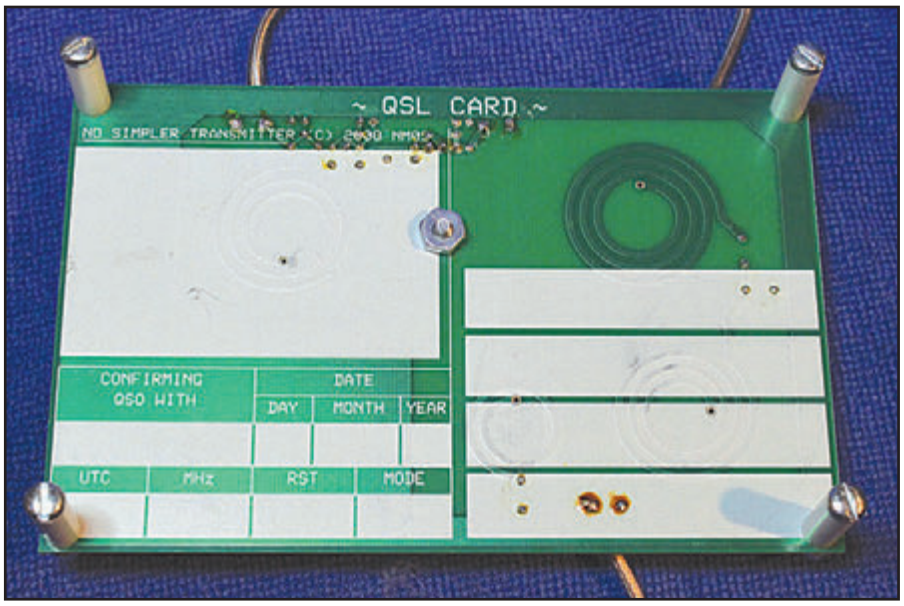
The NS-40 also does not come with any power cables or other connections. Those are easy to do yourself; just be sure that you double check your polarity, especially on the power leads. I advise building the pigtails before assembling the board so that the finishing step of making and attaching the cables goes easily and quickly. When I built my NS-40, I made three pigtails that connect to the board. The first pigtail I made is a connection to a Powerpole connector for 12 VDC. The next cable goes to an inline 1/8-inch jack for connection to either my straight key or keyer output. Finally, I used something I found in a flea market for the RF output. I had found a 4-foot RG-174 Teflon® coax cable that had a male SMA on one end and a female BNC on the other. I cut the female BNC end short to use for the RF output of the NS-40 and used the male SMA end with the longer cable for an RF DF kit connection to my HT. Scrounging this cable in hamfest flea markets, I got two cables for the price of one! Once the board is assembled, the legs for the PC board are attached and the transmitter is ready to go. The NS-40 puts out a full QRP "gallon" at 5 watts and requires no alignment at all.

The NS-40 can be used with either a manual T/R switch or a device such as the 4SQRP Magic Box automatic T/R switch. The NS-40 is on a fixed 40-meter frequency, mine being on 7.040, which is a known QRP watering hole. Of course, being hams, we like to improve on wonderful kits like the NS-40. A great little improvement for it is in the form of another 4SQRP kit, the SAVXO. The SAVXO was designed by Jim Kortge, K8IQY, and like the NS-40, it is sold by the Four States QRP Group.

The SAVXO connects to the NS-40 CW transmitter in place of the crystal and provides a stable 7–8-kHz tuning spread. In addition, the SAVXO itself can put out as much as 250 mw, making it usable as a QRPp transmitter as well! There are

two build options in the manual, one for use as a VXO in most oscillator circuits and the other specifically to drive the 5-watt MOSFET final in the NS-40, which requires more drive. Be sure to carefully follow the part selection for your chosen options, as the components needed for either option are both provided in the kit. If you decide to use the SAVXO as a standalone QRPp transmitter, be sure to download the additional information on the web site for the SAVXO.

Building the SAVXO is pretty straightforward once you have selected your option. Be sure to only install the parts called for during the initial construction before adding your chosen option components. Also, be sure to follow the directions carefully when installing the crystals. A paper insulator needs to be made for each crystal before it is installed in order to prevent shorting it



Bottom of the NS-40a kit featuring the blank QSL card to be filled out with your first QSO.

CommRadio CR-1

The **CommRadio CR-1** is a true SDR, but does *not* require a computer. Enjoy the benefits and performance of state-of-the-art SDR, but in a conventional radio package. The CR-1 SDR is independent of a host PC, using embedded digital signal processing technology that provides a degree of portability and performance previously unavailable to the radio enthusiast. Coverage includes: 500 kHz-30 MHz, 64-260 MHz and 437-468 MHz in AM, SSB, CW, WBFM, NBFM modes. (150-500 kHz with reduced performance). The incredible performance is combined with exceptional portability and ease of use. The radio may be powered via USB or 6-18 VDC input. Enjoy *top-shelf* American technology in a compact, metal case measuring 5.64 x 2.43 x 6.10" 1.8 lbs.

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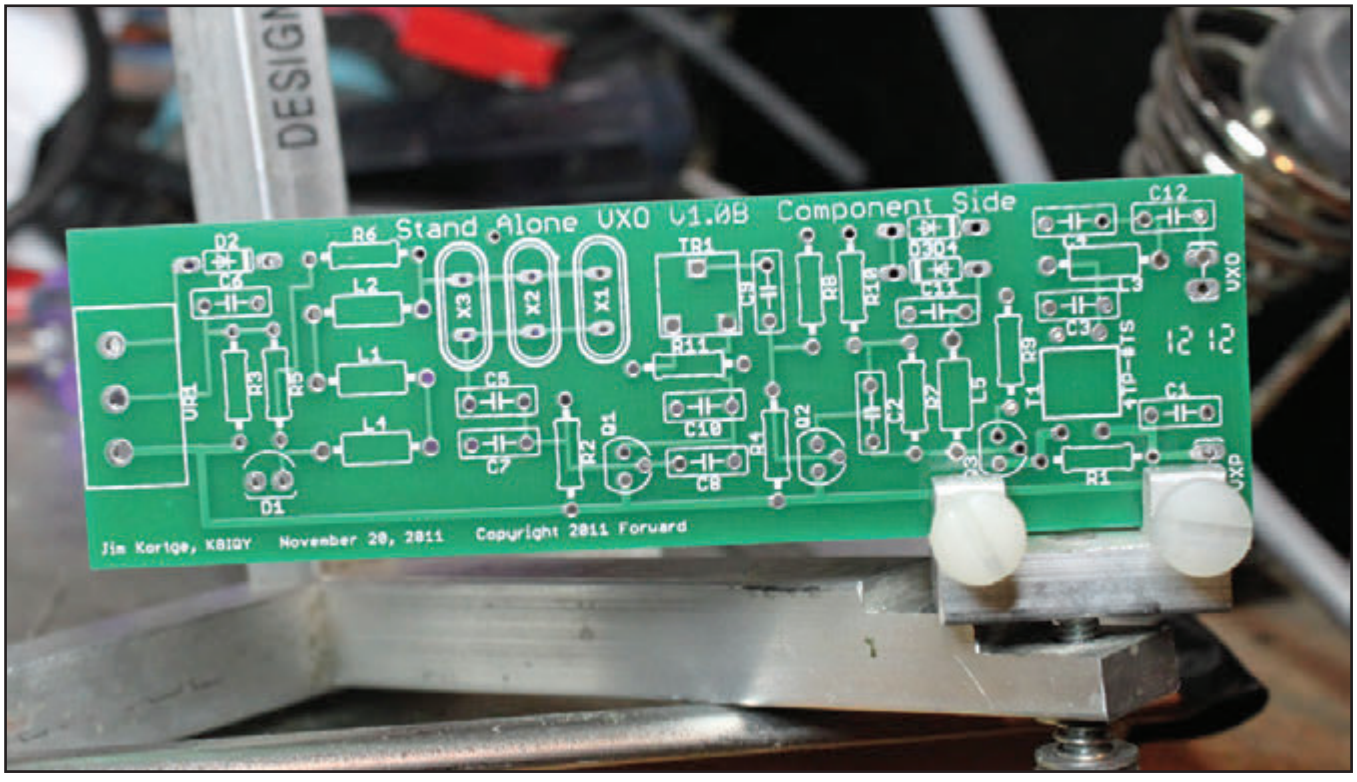
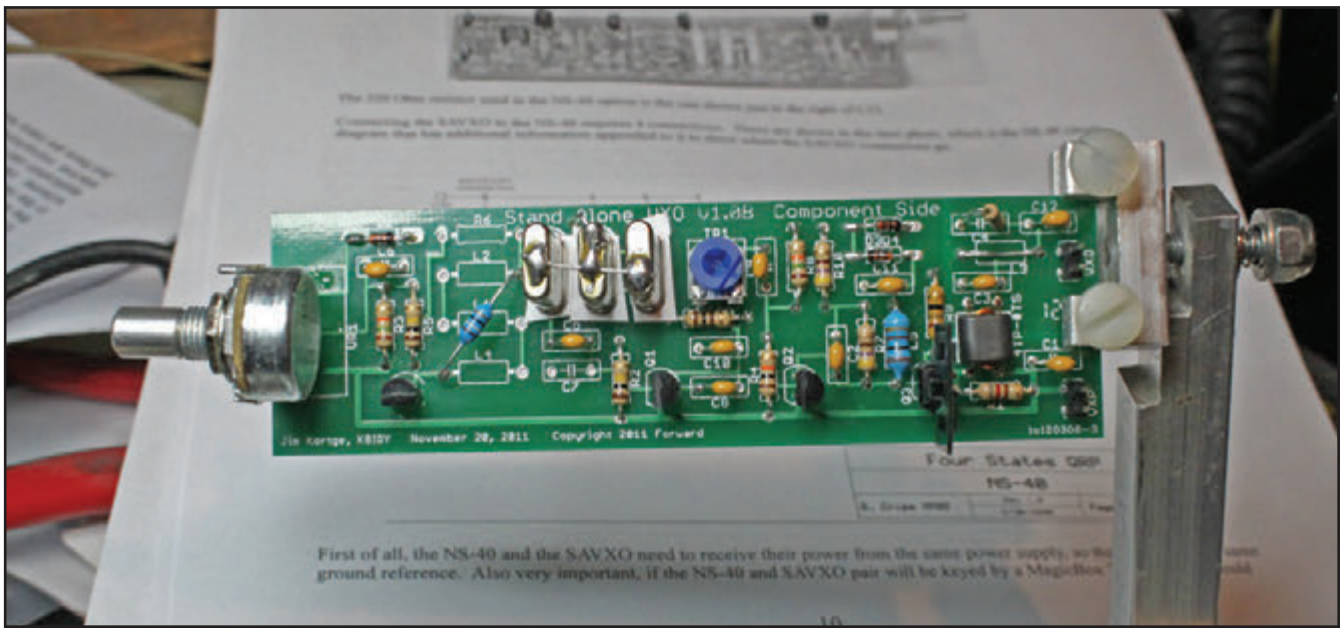


SAVXO board ready for assembly.



Completed SAVXO board with NS-40 option 2 ready to be connected to the NS-40.

what's new

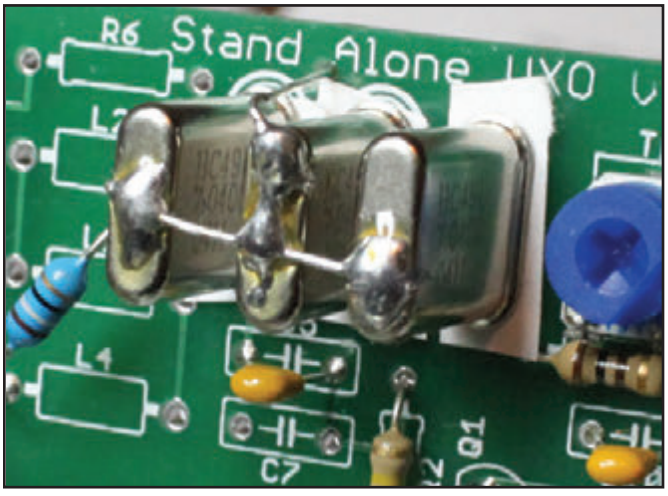
N3FJP Software Amateur Contact Log 4.5

Version 4.5 of N3FJP's Amateur Contact Log program is now available. Features include an update to accommodate Version 2.0 of the ARRL's TQSL software for Logbook of the World, support for the Wireless Institute of Australia's awards program, and a bug fix for interfacing with the Yaesu FT-847 transceiver. Upgrades are free to registered users. You'll find AC Log 4.5 at <http://www.n3fjp.com/aclog.html>; Steps to upgrading are at <http://www.n3fjp.com/upgradestepsaclog.html>.

RF Micro Devices 1.2-GHz Power Doubler Amplifier

Designed primarily for cable TV operators trying to meet new DOCSIS 3.1 cable specifications, the new RFCM3316 Gallium Nitride (GaN) power doubler amplifier from RF Micro Devices (RFMD) may be of interest to amateurs as well. It operates from 40 MHz to 1.2 GHz and has 20% less current consumption than equivalent Gallium Arsenide devices. The company also notes its small footprint as well as ruggedness and excellent linearity. Full-scale production is expected to begin in early 2014. For more information, visit <http://www.rfmd.com/>.

Note: *"What's New" is not a product review and does not constitute a product endorsement by CQ. Information is primarily provided by manufacturers/vendors and has not necessarily been independently verified. The purpose of "What's New" is to inform readers about new products in the marketplace. We encourage you to do additional research on products of interest to you.*



Close-up of the three crystals in the SAVXO with case grounding wires attached.

during installation. There is also a small binocular-type toroid, but it is relatively easy to deal with. Just pay close attention to the number of windings needed for your chosen option. Once completed, a frequency counter or a good receiver is very useful to determine if the SAVXO is operating correctly. Expect to spend around two hours or so putting this kit together.

The NS-40 sells for $32 including shipping in the USA and is sold by the Four State QRP Group at <http://www.4sqrp.com>. The SAVXO is $25, also including USA shipping from 4SQRP.

This probably would be a great time to mention the upcoming Ozarkcon QRP Convention in Branson, Missouri on April 4–5, 2014. This fun weekend features lots of great QRP forums as well as a live hands-on kit building experience! With lots of family things to do in Branson, Ozarkcon is a great family weekend trip. Ozarkcon is at <http://www.ozarkcon.com>. See you in Branson! Until next time . . .

73 de Joe, KØNEB

What You've Told Us…

Our November survey asked about your ham radio-related holiday wish list. Just over half of you who responded either hoped or expected to receive at least one ham-related gift. (Did you get it?) On the other hand, when we asked if you were planning to give *someone else* a ham-related gift, only 9% said yes, 85% said no and 6% weren't sure.

If you could receive only one ham-related gift… 29% of you said you hoped it would be a station accessory, followed by an HF transceiver at 24%, and HF antenna (18%) and a tower to put it on (10%). If you could get *two* ham gifts, then nearly one in five of you each hoped the second one would be either a station accessory or a new/renewed subscription to *CQ* or one of its sister publications (Thanks!). Those were followed by an HF transceiver (15%) and an HF antenna (12%). All of the other options were "in the noise."

Among those of you hoping to get ham equipment of some sort, 56% of you were looking for new commercially-built gear and 32% wanted a new kit to build; and just a handful of you were looking either for parts from which to build something yourself or used gear to operate, modify or repair.

Finally, we asked what broad goal you'd like to accomplish with whatever new gear you got. Forty-nine percent said "enhance my station's capabilities for bands/modes I already use," followed by 19% who looked forward to making operation of their stations easier or more flexible and 13% who wanted to enhance their stations by adding new bands or modes.

This month's free subscription winner is Walt Smith, K1QS, of Naples, Maine.

Reader Survey February 2014

We'd like to know more about you … and especially what's important to you in ham radio and how we at *CQ* can help serve you better. There are two ways to respond to this survey:

*** Respond online** at <www.surveymonkey.com/s/CQFeb14> [From the digital edition, just click on the link].

- OR -

*** Cut out or photocopy** this page

*** Circle the numbers** that correspond to your answers

*** Mail your completed survey to:** November Reader Survey, CQ magazine, 25 Newbridge Rd., Hicksville, NY 11801.

We will continue to select one respondent to each survey to receive a free one-year subscription (or extension) to *CQ*. This month, as we expand our coverage to include more than just ham radio, we'd like to know a little more about your broad radio interests.

1. Which of the following hobby radio activities do you currently enjoy?

Amateur radio / HF 1
Amateur radio / VHF/UHF 2
Broadcast band DXing 3
Broadcast history 4
Personal 2-way radio (e.g., CB, FRS, MURS) 5
Radio technology (including building/restoring equipment) 6
Shortwave listening 7
VHF/UHF scanning/monitoring 8
Other (please specify) 9

2. Which of the following activities provided your *introduction* to the radio hobby?

Amateur radio / HF 10
Amateur radio / VHF/UHF 11
Broadcast band DXing 12
Broadcast history 13
Personal 2-way radio (e.g., CB, FRS, MURS) 14
Radio technology (including building/restoring equipment) 15
Shortwave listening 16
VHF/UHF scanning/monitoring 17
Other (please specify) 18

3. Before this month, to which CQ publication(s) did you subscribe? (Select all that apply)

CQ Amateur Radio 19
CQ VHF 20
Popular Communications 21
WorldRadio Online 22
None 23

4. What is your primary choice of device for reading a digital magazine?

Desktop computer 24
Laptop computer 25
Tablet (e.g. iPad) 26
E-reader (e.g., Kindle Fire) 27
Smartphone 28
None 29

Thank you for your responses. We'll be back with more questions in upcoming issues.

Survey Response for Issue:

Name _____ **Call Sign** _____

Address _____

City _____ **St/Prov** _____ **Zip/PC** _____

Country _____

E-mail _____

BY DON ROTOLO,* N2IRZ

How to Create an Inexpensive Software-Defined Receiver

It has been a while since we got into some deep digital, but with the new year we are changing that. This month we'll see how to create a very cheap (less than $15) software defined receiver with a frequency range of about 50 MHz to 2.2 GHz. What would you do with such a receiver?

The business end of this receiver is a USB TV reception dongle used to receive digital video in much of the world (but not the USA and Canada). In normal use, the receiver just down-converts the received signal and delivers I and Q data to the computer for TV decoding. (I and Q are the In-phase and Quadrature [90° phase] amplitude-modulated sinusoidal components of an angle-modulated signal. Reducing an angle-modulated signal [such as FM or PSK] to two AM signals greatly simplifies the post-processing of the signal in the digital realm.)

I found several suppliers for these receivers on eBay and ended up buying one from Hong Kong for $12.97 including shipping. It did take almost three weeks for it to arrive, but I was in no particular hurry. You can buy one from a USA-based seller for a somewhat higher price if time is a consideration.

What you want to buy is a DVB-T (Digital Video Broadcasting – Terrestrial) USB dongle that uses the Realtek 2832U control chip and the Elonics E4000 tuner chip. There are dongles that use other tuner chips, such as the FC0013 (22–1100 MHz) or the R820T (24–1766 MHz), and these will work but with a different tuning range. The E4000 has a specified tuning range of 64 to 1700 MHz, but it can be operated outside that range if we can accept less-than-optimal LNA (low noise amplifier) performance. Note that there is a gap in frequency coverage on the E400 from about 1100 to 1250 MHz.

To control this receiver I used a program called SDR# (pronounced "SDR Sharp"; see <http://sdrsharp.com/>). For me, the attraction of this particular program is that it runs under Windows®. If you have a Mac or Linux machine, you have other options such as Gqrx <http://gqrx.dk/>, or search the Internet for several other choices. SDR# can decode several common modes (SSB, FM, AM, and CW) using your computer's sound card and offers a familiar waterfall display.

If you find the frequency range of 50 MHz to 2.2 GHz somewhat limiting—I'm thinking of the HF bands here—you can buy an upconverter to shift the received frequencies up to where we can use the SDR to receive them. NooElec <http://www.nooelec.com> sells its Ham It Up v1.2 upconverter for $43, and it shifts received signals up in frequency by 125 MHz to allow signals from almost DC to about 65 MHz to be received. I'd wanted to buy one to try it out, but that wasn't in the cards; however, a review of the documentation on its web site has convinced me that it is easy to install and use. If you have or buy one, let me know what you find.

Of course, a receiver is only as good as its antenna. Outside the specified frequency range of 64 to 1700 MHz the receiver loses some of its sensitivity, making a good antenna that much more important. My SDR dongle came with an antenna, but for all but the most powerful broadcast-band stations it was useless. Its only real value is the connector and coaxial cable, on which I mounted a BNC connector, simplifying connections to my various antennas for testing. (See fig. 1.)

My outdoor ground-plane antennas for 2 meters and 70 cm worked well enough for the local repeaters, as did my loop Yagi on 23 cm. I was also able to use the 2-meter antenna to pick up some nearby aircraft on the local FAA approach frequency, as well as the town police and fire departments.

For the more adventurous among us, consider a low-noise preamplifier for the band(s) in question from Down East Microwave (http://www.downeastmicrowave.com/) or the PR-2 broadband

Fig. 1. The USB Dongle and unmodified antenna. The dime is there for scale. This particular unit receives from 52 to 2202 MHz, with a gap between 1102 and 12324 MHz. The antenna is only useful for the end connector that mates to the Dongle.

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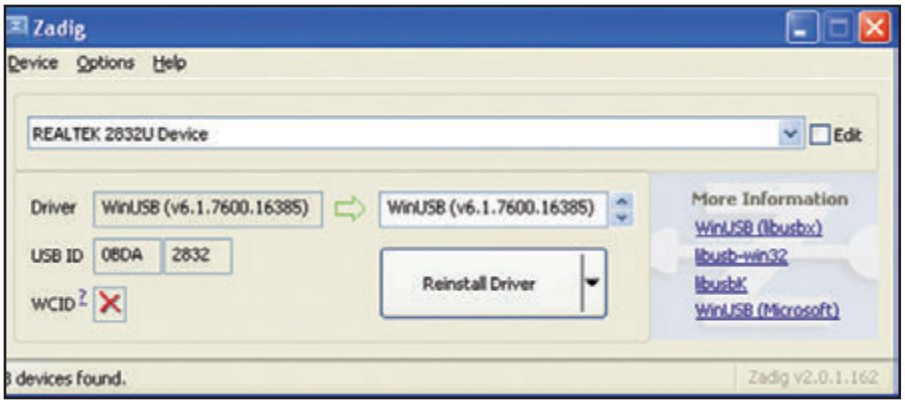

Fig. 2. The Zadig driver installer for Windows® XP. The WinUSB driver is absolutely essential to getting SDR# to recognize the dongle, and for me was by far the most difficult challenge in getting everything to work. Just make sure you install the driver for the USB Dongle and not for some other device! Check the USB ID as explained in the text to make sure.

preamp sold by Ramsey Electronics (http://www.ramseyelectronics.com/).

Hardware

The first step in making all this happen is to buy the DVB-T USB dongle. As I stated before, you want one with the Realtek RTL2832U control chip. The preferred tuner is the Elonics E4000, but any tuner will work, just with a different frequency range. On eBay, I did find several listings that mention the E4000, but upon careful reading I found the device they were offering had a different tuner chip in it. Some were claiming that this or that chip was an "Upgrade to the E4000," but don't believe it. Do your research, and buyer beware.

Once the unit arrives, the first step is to throw out the software disk that comes with it (although mine just came with a web address). If you happen to install the software and driver, I am told you'll have a devil of a time getting rid of it.

Software Installation

Visit the SDR# downloads page at <http://sdrsharp.com/index.php/downloads> and then find and download the file "sdr-install.zip." Unzip it to a directory in My Documents, and run the file "install.bat." This will automatically go to various websites, download the necessary files, and place them into a directory named "sdrsharp." You can go and download things manually, following the instructions in the *SDR# Quick Start manual*, but this batch file really does work nicely and eases installation considerably.

Note that by "installing" all these files to a directory, SDR# is not really "installed," but merely copied. This means that the Windows® registry is not touched, and you can move the entire installation from one computer to another simply by copying the files.

Zadig Driver

Once you have the USB Dongle, connect it to a handy USB port on your computer. With some luck, Windows® will recognize the device and install some kind of driver; allow it to search the Internet if it asks. It doesn't really matter which driver it installs, since we're going to replace it, but having a driver makes the next step easier. It took my computer a few tries before it installed a driver (and even then it took several minutes), so be patient and persistent.

Once a driver installs, you will want to *replace it with the Zadig driver.* The version of Zadig that is installed by the batch file is an older version that should work fine in Windows® XP or later, but if you download it yourself from the Zadig website <http://zadig.akeo.ie/> as I did, make sure you get the correct file for your version of Windows®.

This is a summary of the instructions that can be found at <http://rtlsdr.org/softwarewindows>, along with some of my own observations. Run Zadig.exe and, after you allow the program to run, you'll see the window shown in fig. 2. Select "Options" and check "List all devices." Then click on the devices list and pick the entry that says "Bulk-In, Interface (Interface 0)." If your list doesn't have this entry, don't panic: You instead may see an entry such as "RTL 2832" or something very similar. Mine said "REALTEK 2832U DEVICE" as shown in the image. To be certain, check the USB ID numbers against the hardware list at <http://

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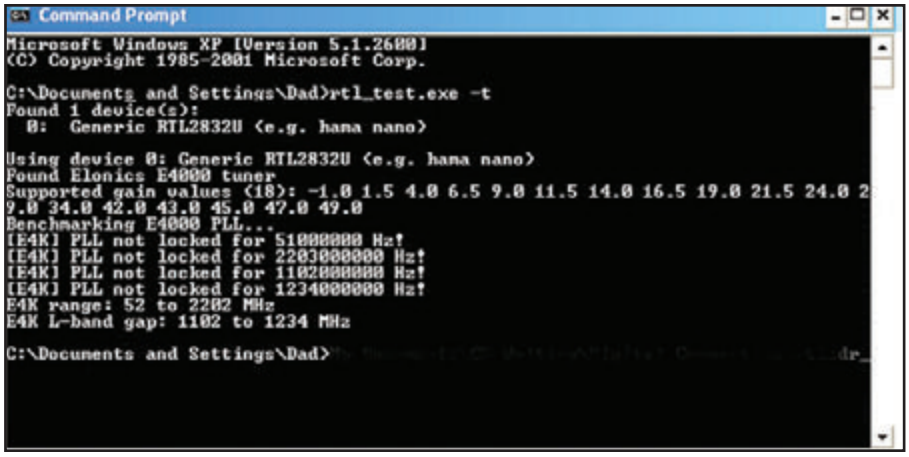

Fig. 3. Running rtl_test.exe –t in a Command prompt window verified my device was working properly, as well as telling me the expected bandwidth and gap, and verifying that I have a real E4000 tuner installed and not a less-capable "equivalent."

rtlsdr.org/hardware-usb>, and if yours are listed, you are on the correct entry. WARNING: Being on the *incorrect* entry can cause a problem. You are replacing the original driver, so whatever is using that driver (such as a keyboard) probably will stop working.

OK, so whatever the device name is, if you've verified the USB ID is on the hardware list, you should make sure you're going to install the *WinUSB* driver (and not the *libusb* driver). Go ahead and click the "Install driver" (or Reinstall Driver—the button text changes depending on what's in there) button. Installation should take about a minute.

Now that we've installed the driver, start SDR# by double-clicking the SDRSharp.exe file. (You might want to right-click it, and pick "Send to ... Desktop" to create a shortcut). You should see a display something like fig. 3.

Still a Few Bugs in the System

In theory, if you select the "RTL-SDR / USB" device at the upper left and set everything correctly (Select Wideband FM, connect a decent antenna, tune to a big local FM station, and click "Play") you'll be receiving signals. Ah, but sometimes reality gets in the way of theory. While most users report a completely problem-free installation, I was one of the lucky ones who had to spend a day fooling with the driver.

Oh, it was installed, but SDR# alternately didn't recognize the driver and the device, sometimes complaining the "RTL-SDR/USB is either not connected or its driver is not working properly," and other times complaining "RTL-SDR

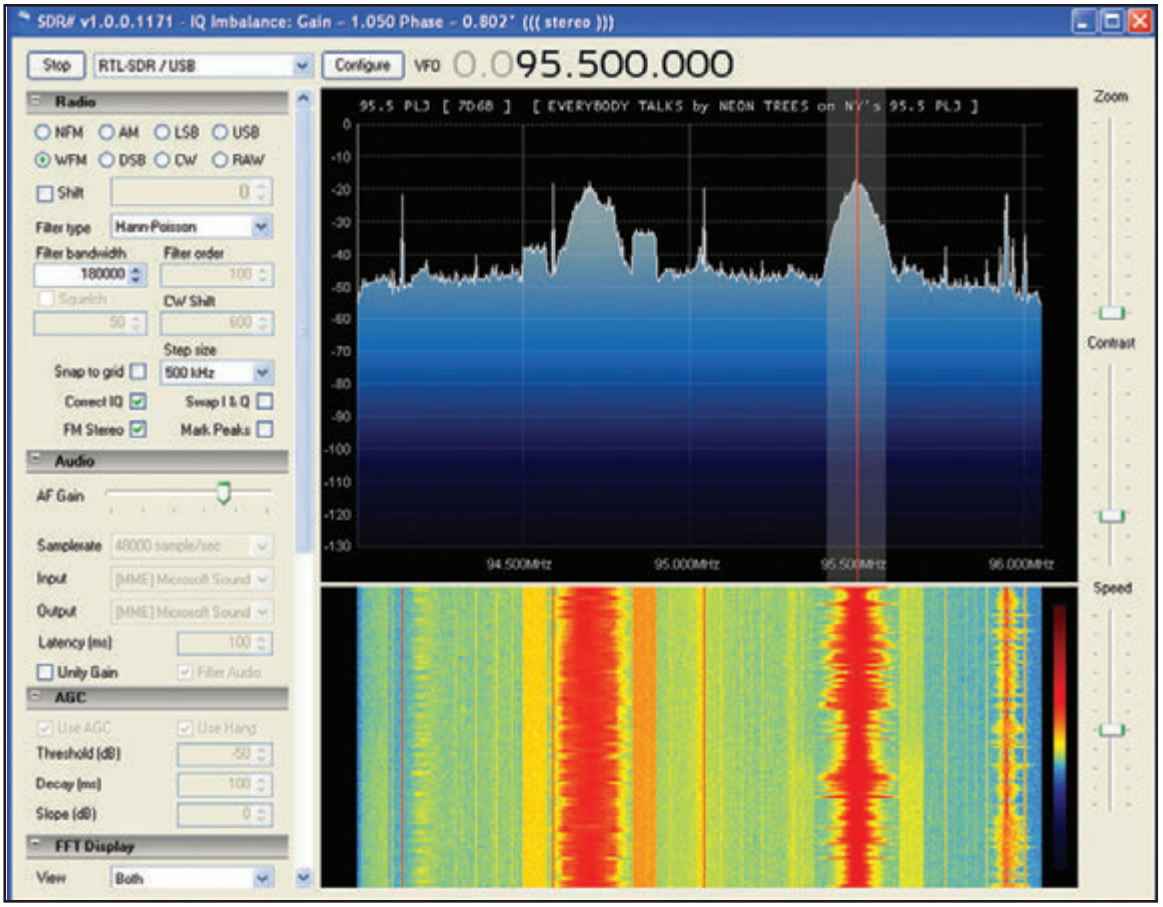

Fig. 4. SDR# tuned to New York FM station WPLJ, 95.5 MHz. Note the station's crawler text (along the top of the spectrum window). The station at 94.7 MHz has clearly visible HD Radio sidebands. I have SDR# configured for Wideband FM, with a filter bandwidth of 180 kHz, FM Stereo decoding enabled, and SDR#'s exclusive I/Q correction system, which helps correct any I/Q imbalances, also enabled.

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USB device not found." I reinstalled the Zadig driver several times, also trying the "libusbK" driver, rebooting each time—maybe 25 times over the course of a whole day. In between tries, I was scouring the web looking for help. I found the USB Dongle itself was working normally by running "rtl_test.exe" with the "-t" option (found within the ZIP file at <http://bit.ly/HRswcR>); see fig. 3. I did manage to actually crash my PC once, and also had to reinstall SDR# twice. It would work a little sometimes, and not at all other times, including one time with no reception at all.

Ultimately, I cannot really say what happened to make it work, but it did actually work—and that's when I found out how important a decent antenna is. Even when it was working well, sometimes it would stop tuning, or stop receiving any signals, and this was usually cured by clicking "Stop" and then "Play" again, but sometimes I had to close and restart the software for it to come back. I suspect that a flaky USB connector might have had some effect on this.

I also forgot to mind the band edges, trying to tune in WABC-AM on 770 kHz, a very powerful local AM broadcaster. I noticed that one pretty quickly. I did manage to get most of the commercial FM band, where the HD Radio subcarriers are clearly visible (fig. 4), and SDR# even showed the low-speed text crawler that some stations have. I then monitored a local 2-meter repeater and found that the default 10-kHz filter bandwidth for Narrow FM was way too wide: Setting this around 5 kHz dramatically improved the received audio.

Fun with the Receiver

I've been having fun playing with this receiver. The software is easy to learn, although I did refer to the Quick Start Guide a few times for an explanation of some of the settings. The receiver itself is as small as a postage stamp, and almost as light, making it perfect for travel, a spare receiver for field work, and handy to have for emergencies: During superstorm Sandy last year, broadcast radio was about the only source for news and information, since phones, power, and internet were out almost everywhere.

3D Printers

Since my pair of columns on 3D Printers went out (October and December 2013 issues), the Oak Ridge National Laboratory, the Department of Energy's Advanced Manufacturing Office, America Makes (formerly NAMII), MakerBot, TNFIRST, 3D Systems, and MakerGear decided to make a 3D printer available to thousands of high schools around the world that have a FIRST Robotics Team <http://www.usfirst.org/>. They are starting with 400 printers this year and expanding it to most teams next year. Although we're not quite at the point where 3D printers are able to produce almost anything, we're not too far from that either.

A colleague compared the current state of the art in 3D printers to the state of the art in home computer printers in 1990. Back then, I had a 9-pin dot-matrix printer that was slow, noisy, jammed often, and used special paper with tractor feed holes along the sides. How primitive those were in comparison to what we can get today! In just a few years, 3D printers that can produce real and useful items of many materials will be available at Walmart for $29, but surely the "ink" will cost twice that! As for me, *I just can't wait.*

That's all this month, and I'll be back in this spot for the April edition. In the meantime, give the USB receiver in this month's column a try, and write to let me know how you made out, especially with that driver installation.

Until then . . . 73 de Don, N2IRZ

learning curve

BY RON OCHU,* KOØZ

Repeaters

In a previous edition of this column I wrote that a repeater serves as a club cornerstone. Many of us use repeaters and we want to "pay our fair share" by supporting the club sponsoring the repeater. Repeaters serve a vital ham radio function, for they allow us to communicate with each other with lower power radios and vertical antennas over a greater distance than would be possible without them (fig. 1). I think most hams are familiar with a repeater and have even used one, but how many hams have actually seen a repeater, much less be able to identify its specific parts?

Location, Location, Location!

Well, if you wanted to learn more about repeaters then today is your lucky day; read on, dear reader! I would like to thank Skip Riba, WS9V, for taking me on a tour of his UHF repeater (photo A). Skip's repeater is located on a 480-foot tower just northwest of Litchfield, Illinois (photo B). This is farmland and prairie country. The land around here is flat for miles. When I moved here about two years ago, I asked Skip if he knew of any high spots in the county and his response was "If I did know of any, my repeater would be on it!" Hence, his repeater is located on a former Western Union microwave relay site. From this location, Skip's repeater has consistently good coverage throughout a 60-mile diameter circle and fairly good coverage over a 100-mile diameter! Add icing to the cake, and Skip linked his repeater with ones in five nearby cities (fig. 2), giving hams in the area seamless FM coverage throughout most of central and southern Illinois!

In the VHF/UHF world height is everything. That's why out in the mountain states, repeaters are located on mountain peaks. Here in the Midwest, water towers, tall buildings, or former commercial microwave relay sites are sought after for their height above average terrain (HAAT). Unfortunately, these sites are becoming increasingly hard to find because of tightened security and commercial competition. Cell-phone companies want these sites, too, and they are willing to pay top dollar to rent antenna space, which is putting these sites well beyond the budgets of most clubs. Sometimes stressing the public-service aspect of amateur radio is one way to help secure antenna space at reduced cost on municipal government water towers or tall hospital buildings. The WS9V repeater system is directly linked

*e-mail: <ko0z@cq-amateur-radio.com>

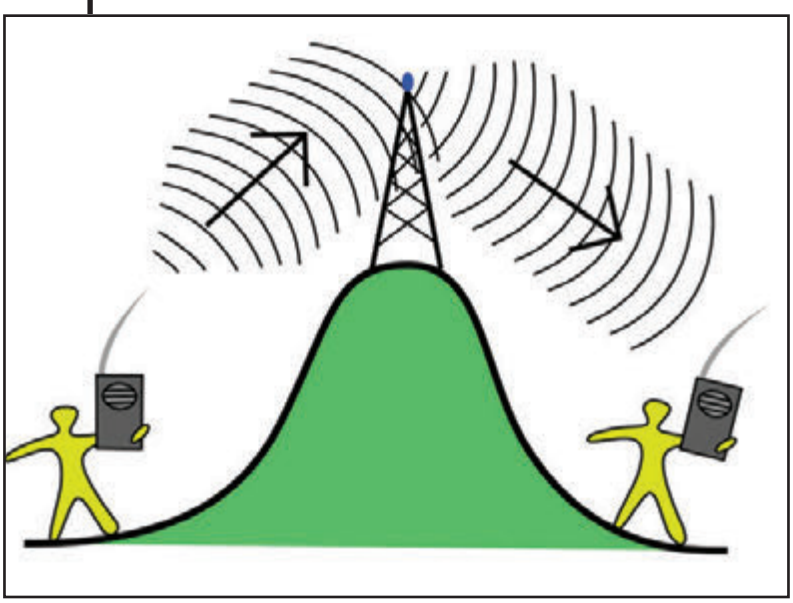

Fig. 1. Repeaters extend the range for HTs and mobile rigs. (Figure by Mouagip, via Wikimedia Commons)

Photo A. Skip Riba, WS9V, standing next to the mighty 444.250 linked repeater serving central and southern Illinois and providing a direct link to the St. Louis National Weather Service office. (Photos by the author)

Photo B. The WS9V repeater is at a former commercial microwave relay site near Litchfield, IL.

Photo C. An example of a commercial VHF 4-bay (folded-dipole configuration) repeater antenna.

back to the National Weather Service St. Louis office, and it provides an invaluable SKYWARN® link for that office's eastern counties watch area.

Let's Take a Look Under the Hood!

Once a high spot for the repeater is secured, the first thing to think about is putting up a durable commercial antenna. These antennas can withstand the harsh environment of a repeater site. It's no fun to replace a repeater antenna that weighs 24 lbs and is 20 feet in length, especially so at the WS9V/R site, which requires climbing a 480-foot tower clear to the top. Another feature of a commercially built antenna (photo C) is its radiation pattern. Besides being omnidirectional, it also can focus most of its energy out toward the horizon, which maximizes the gain (ability to transmit and to receive signals over greater distances). However, these antennas are not inexpensive. With shipping, they can cost nearly $1000, and many site owners require that you hire professional installers to put them up.

Minimizing Feedline Losses

Having a high-gain antenna is a must for a repeater site, but it is also vitally important to minimize feedline losses. This means that hardline coax is utilized to carry the signals between the antenna and the repeater (photo D). At WS9V/R, Skip uses Andrews 1-1/4-inch hardline to minimize signal losses. Again, this is a significant expense. New Andrew 1-1/4-inch hardline costs over $10 a foot. Connectors for each end are well over a $100 each.

Why use such large diameter hardline? The published specs for $1^1/4$-inch hardline at 450 MHz show a predicted loss of .55 dB per 100 feet. Now compare that figure with that of high-quality LMR 400 low-loss coax, which at 450 MHz has a rated loss figure of 2.7 dB per 100 feet. With a 500-foot run to the antenna, the hardline has approximately 3 dB of loss, or about half the power that you started with. In the same situation, the LMR 400 would suffer losses of 13.5 dB! Keep in mind that every 3 dB either adds to the power by a factor of 2 (gain) or decreases it by one-half (loss). Thus, it's clear why you want to minimize feedline loss in this situation. Adding the connectors introduces some slight signal loss as well; therefore, good-quality "N" connectors are a must.

Are We There Yet?

Okay, we now have the RF from the antenna going down the tower leg and almost into the shack. Prudence and experience teach us to make sure that the hardline and the antenna are at a DC ground. Photo E shows a grounding strap from the hardline to the tower leg, which in turn goes to an earth grounding rod. Skip has his hardline grounded near the

antenna and halfway down the tower as well as at the bottom of the tower, and all of that goes to a common-point ground on a grounded copper plate inside the building. A quick-acting gas-filled cartridge lightning protector such as a Polyphaser is also part of the grounding system. Now we are ready to introduce the antenna system to the repeater system!

What Else Makes Up a Repeater?

A repeater typically consists of five major components in addition to the antenna and feedline (fig. 3 and photo F): a receiver, a duplexer, a controller, a transmitter/amplifier, and a heavy-duty power supply. Let's take a look at each of these elements and how they work together.

For starters, most repeaters listen on one frequency and transmit on another. The WS9V repeater system listens on 449.250 MHz and transmits on 444.250 MHz. Consequently, when I program my mobile FM transceiver or my HT, I need to include the proper "offset." By that I mean that when I dial in 444.250 MHz my radio is *listening* on that frequency, but when I press the push-to-talk (PTT) switch, my radio now trans-

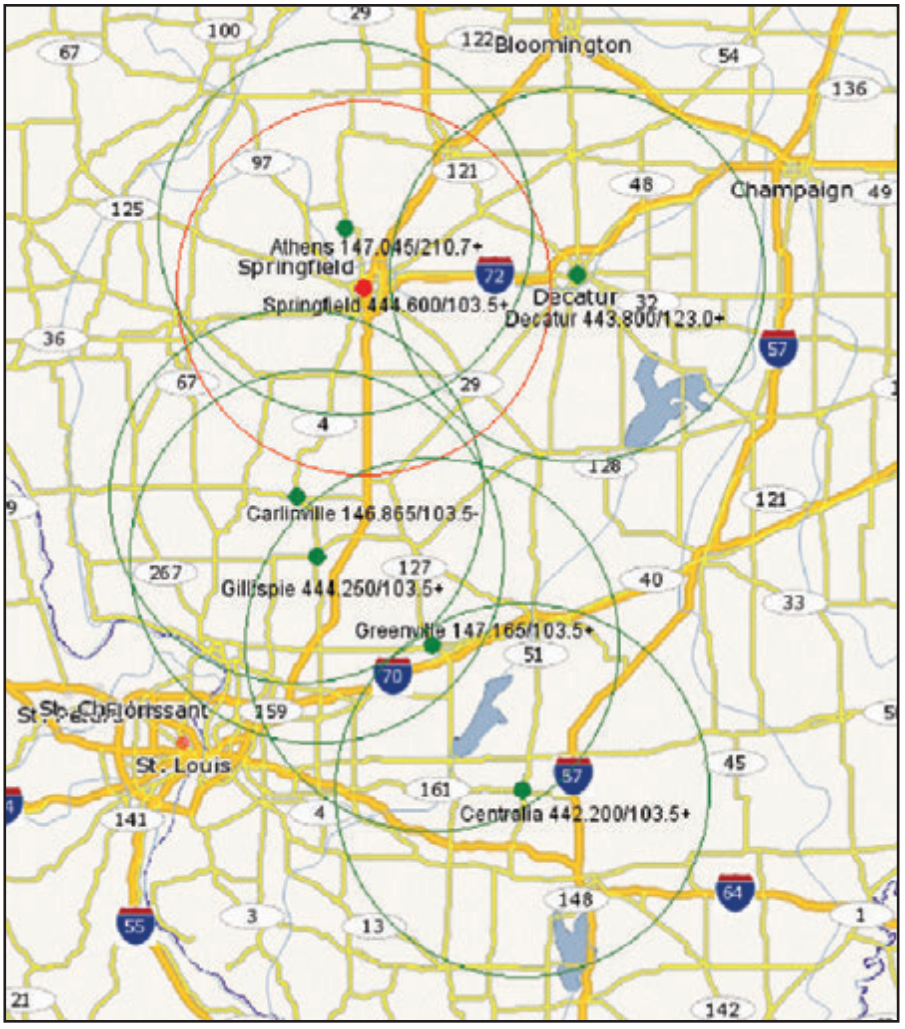

Fig. 2. WS9V linked repeater service area serving central and southern Illinois. (Used with permission, from the K9MCE.org [Macoupin County ARC] website)

Photo D. Some examples of 1/2-inch hardline (left) and 7/8-inch hardline (right). WS9V uses 1 1/4-inch hardline in his repeater system.

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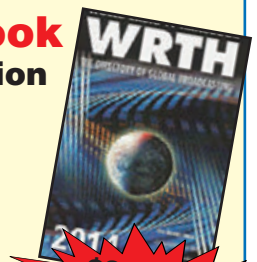

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mits 5 MHz higher, at 449.250 MHz, which, not coincidentally, is the frequency on which Skip's repeater is listening! (A 5-MHz "split" is standard on 70 centimeters, or 440 MHz.)

By convention, hams name a repeater by its transmit frequency (your receive frequency). For example, "I am going to QSY to the 444.250 repeater." When I am in the St. Charles area, I use the 146.67-MHz WBØHSI repeater. That repeater transmits on 146.670 MHz and listens on 146.070 MHz. Notice that these frequencies are only 600 kHz apart, the standard for the smaller 2-meter band. It is important to program your radio with the proper offset. Current FM transceivers and HTs have automatic offset switching for each ham radio FM band.

Duplexer

Duplex means to transmit on one frequency and to receive on another. Most repeaters are "in-band," which means transmit and receive frequencies are within the same band, such as the WS9V repeater, which is on 70 cm. A duplexer allows a repeater to use only one antenna and one feedline, even though it is receiving and transmitting simultaneously on the same band. The other option is to have a separate transmit antenna with feedline along with a

Photo E. A good ground is a must for lightning protection and lowering noise levels.

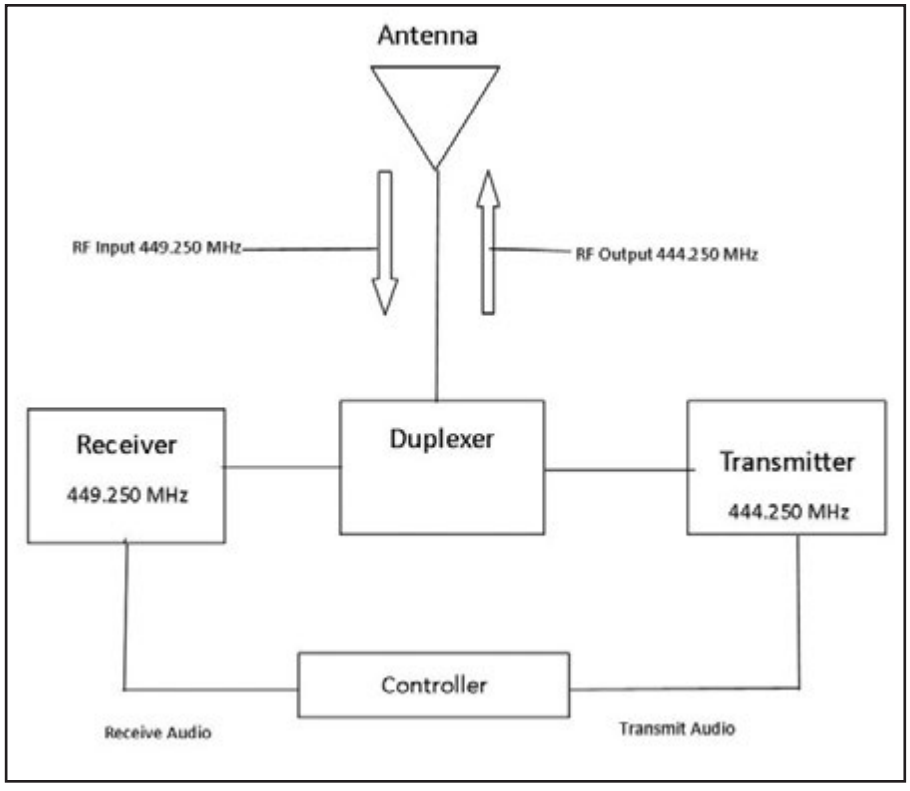

Fig. 3. A block diagram of a basic repeater system.

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Photo F. The front side of the rack-mounted WS9V repeater system. On the top is the 449.250-MHz receiver and below that are the ACC RC-96 controller, the 444.250 MHz transmitter, the 70-cm BpBr duplexers, the TPL 180 watt amplifier, and the Astron 70 amp 13.8-VDC power supply. Not pictured is the battery back-up in case of commercial power failure.

separate receive antenna with its own feedline. If both antennas are going to be at the same site, then they need to be physically separated from each other—either vertically or horizontally—by a calculated wavelength distance, which greatly complicates the repeater's mechanical arrangement (not to mention the cost). Duplexers, on the other hand, make a repeater installation a whole lot simpler.

The WS9V repeater uses a four-cavity Wacom BpBr duplexer (photo G). Bp means *band pass* and Br means *band reject*. In essence, two cavities are connected to the repeater's transmitter/amplifier and they allow only the transmitter frequency of 444.250 MHz to pass on out to the repeater antenna. The other set of two cavities rejects all other frequencies and allows only signals on 449.250 MHz to pass through to the repeater receiver. It does this simultaneously using only one antenna and hardline for both transmit and receive! Like everything else, duplexers are not inexpensive. A vital part of a duplexer is to have low-loss coax cable connecting it to the receiver and the transmitter. This prevents unwanted noise and interference from getting into the system.

Receiver

Skip's repeater uses an old Spectrum Communications UHF receiver. Although this receiver is old, it still has high sensitivity and selectivity. After all these years, it keeps on working 24 hours a day and seven days a week. The audio from the receiver is now fed to the repeater controller through low-loss coax.

The Repeater Controller

The controller is the "brains" behind a repeater. It is a computer circuit that controls a number of functions such as when to come on or off. It has a built-in timer that is set to limit the amount of time the repeater transmits before "timing out," or automatically shutting down the transmitter. Most repeaters try to keep a person from being "long-winded" by timing out before three minutes have elapsed since the repeater was first keyed-up. That's why it is proper repeater etiquette to keep your transmissions short and to let the repeater reset its timer. You'll usually know when the timer is reset by the courtesy beep.

If the repeater is set up with tone access, the controller takes care of that function as well. Many repeater owners require the use of a CTCSS (Continuous Tone Coded Squelch System) tone—also frequently referred to by its Motorola trademark of PL, or Private Line—to access the repeater (WS9V/R uses a CTCSS tone of 103.5 Hz). This is a subaudible tone that is transmitted on your RF carrier, which tells a repeater's receiver to allow your signal to be retransmitted. This system goes a long way toward reducing a lot of noise and interference during band openings.

The controller also has a carrier operated relay which senses an incoming signal and then sets all of the programmed repeater control functions into play. There is also a port for a phone line and a command receiver (222 MHz or higher) for control operator functions such as disabling the repeater, putting it into "weather mode," etc. Finally, the controller identifies the repeater with a voice or CW callsign identification. It now takes the receiver audio and passes it on to the transmitter/amplifier combination.

Transmitter/Amplifier

Once the controller sends the audio to the transmitter, it is now able to take the received audio and pass it on to the repeater's amplifier. The WS9V re-

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peater uses a Spectrum transmitter and its RF output feeds a TPL UHF amplifier which amplifies the transmitter's RF to 180 watts. The transmitter and amplifier now send the received 449.250-MHz signal—at a much higher power level—back up to the repeater antenna on 444.250 MHz. This repeat function is exactly what greatly extends the range of an HT or a mobile radio! By using this repeater you have the equivalent of a 180-watt transmitter with an antenna at 480 feet!

Repeater Trustees are Rock Stars!

Hopefully this little tutorial has given you a better understanding of what goes on with the inner workings of a repeater system. When a repeater is up and running smoothly, almost everyone is happy, except for those poor chaps with nothing but an HT and an inefficient rubber-duck antenna in a low spot 20 miles from the repeater with hill in between, who complain that the repeater isn't working correctly. Perhaps this column will cause them to take pause and to think before speaking. Also, if perchance the repeater system ever malfunctions (and it will over time), don't be too quick to criticize the folks maintaining the repeater.

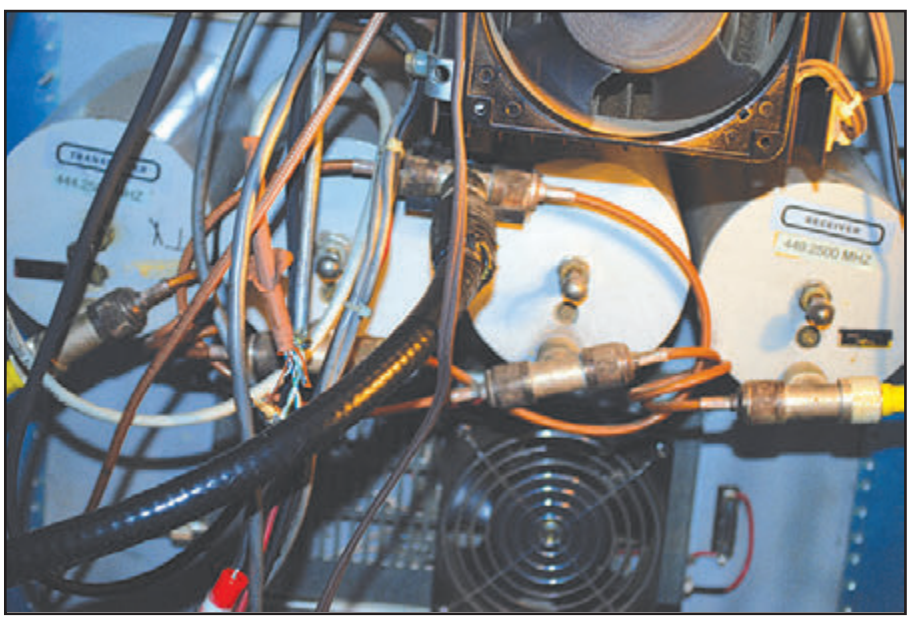

Photo G. These are the 70-centimeter duplexer "cavities," one pair each tuned to the WS9V input and output frequencies.

As you know now, there's a lot of work and a lot of expense in keeping up a system that we all take too much for granted. Therefore, the next chance you get and why not take a few moments to thank the repeater trustee for all that is done to make your daily work commute less boring and something that you even look forward to while in QSO with your buddies. That goes double when the repeater is needed during times of emergency. Please support all this work by joining the group that sponsors the repeater. 73, Ron, KOØZ

BY JEFF REINHARDT,* AA6JR

No Holes in Your Vehicle, Cold-Weather Concerns, and Rovers

Thanks to the many *CQ* readers who responded to the call for photos, ham radio license plates for "The Wall," and for sharing their experiences installing and operating gear for mobile operations. Your contributions make this column more enjoyable—and beneficial—for others.

If you want to check out the growing photo wall, it's found at: <http://www.flickr.com/photos/cqphotogallery/sets>. We're doing pretty well on our quest for WAS with license plates but have yet to hear from some pretty prominent states. Is yours represented? If not, send a photo of your plate to me at the address shown on the first page of this column.

*5904 Lake Lindero Drive, Agoura Hills, CA 91301
e-mail: <aa6jr@cq-amateur-radio.com>

Photo 1. Simple installation of magnetic mount dual-band on the wheel well.

No Holes?

Many hams express concern at the notion of drilling holes in their car or truck, with good reasons. An interior hole can damage whatever lies behind the panel being drilled, be it wiring, a computer module, fuel line, sensor, or other device. In some hybrid or all-electric cars, a very powerful battery or high-current cables can also be present! Of course, too, exterior holes, if not properly done, can damage paint, bring on the beginnings of corrosion, and diminish the vehicle's value when it's time to trade or sell—unless of course you sell it to another ham who sees the utility of your installation.

Therefore, we'll share a very thorough story submitted by *CQ* reader Grady E. McCright W5GEM, who sent the following:

"Several years ago, I splurged and purchased a 2008 Ford F-250, Super Duty, mostly to pull my travel trailer but also to satisfy a long-held desire to own one. Immediately, I was confronted with the installation of a mobile VHF/UHF dual-band transceiver. I had no desire to mar the appearance of my new pickup, so I endeavored to generate a no-holes installation that would both serve me well and meet my criteria of punching no new holes in the vehicle.

"I decided to make do with a magnetic mount, so I purchased a popular-brand dual-band antenna. First I installed the antenna in the bed of the pickup for testing, and I just placed the strong magnetic base on the left-side wheel well (photo 1). The best location, of course, would be at the center of mass and as high as possible. However, since my car port is too low to accommodate such an installation, and since I, without thinking, purchased an aluminum tool box, I would have to crawl into the bed and put the antenna on the roof after taking off the cover each I used the vehicle, and then remove the antenna again before parking in the carport. Therefore, to avoid that tiring issue and after testing the antenna with an antenna analyzer, I chose the wheel well as the permanent location since it tested in the more than acceptable range. The best of both worlds would be leaving the feed line long enough to accommodate the roof installation and simply store the antenna on the wheel well. The roof would be preferred for longer trips and serious operation, yet the wheel-well location would work for casual local contacts.

"I continued with my no-holes installation by routing the feed line through an existing penetration in the front corner of the bed. That access hole was covered by an easily removable rubber plug which

Photo 2. Closed ashtray door that is slanted toward the driver.

Photo 3. Control head lying on the ashtray door with the door spring holding it in place.

Photo 4. Microphone in its nesting place under the seat. The power wires can be seen behind the mic going to the main body.

I saved for future re-installation. Although there are many ways to enter the cab with the coax, I chose to route mine along the frame to the engine compartment using plastic flexible ties to secure the cable. I then penetrated the firewall through the rubber seal around the steering column.

"There is another easy way for the feed-line cable to enter the cab by going through the small air vent behind the rear seat. These small vents allow a small amount of air exchange with the outside and also to permit the doors to close without compressing the air inside the cab. I have used this path on fire trucks in the past and it works fine. Nevertheless, since my truck is a super cab (once called a crew cab), I felt it easier to go through the firewall and not be required to route the cables under the floor carpet or mat to the front-seat area.

"Transceiver power is routed through a fuse in each leg directly to the battery via the same penetration along the steering column through which the coax goes. My dual-band transceiver is an ICOM IC-208 that has a detachable control head, so the body of the transceiver is installed under the middle of the front seat resting on a Sticky Pad (available from automotive stores and used to temporarily hold cell phones, GPSs, etc}. I laid the Sticky Pad on the carpet and then simply set the transceiver body on the top. This installation has been over some really rough roads in the western United States and I have had no trouble with movement.

"The coax cable, the power cord, as well as the remote-control-head interface cable all are routed from under the front of the rubber mats to the main body location beneath the front seat. Lying on top of the carpet and under the rubber mats has made a neat and serviceable installation. The route of the cables is from the center of the floor to keep the cables away from the driver's feet.

"My control head is a perfect fit in the ashtray, I simply opened the ashtray door and placed the control head on the drop-down door where the door spring holds the head securely. It does not move during operation or travel. It is canted toward the driver's position and is effortlessly visible, easily assessable, and convenient. (See the closed ashtray door in photo 2 and the control head installed in photo 3.)

"While one could use an adhesive-backed microphone holder in a suitable location on the dash, I really did not want that nor did I find a good position on the dash, and I could not locate anything metallic on the surface of the

Photos 5 & 6. Kentucky reader Wayne Whitman, K4ELO, sent along these photos of his very clean-looking installation on his pickup truck.

dash. Therefore, I took a magnetic mic holder and placed it on the driver's seat frame just under my right leg (photo 4). When in use, I place the microphone in the seat under the center armrest. Since the microphone is stored and used in close proximity to the main body of the rig, I plug the mic cable into the jack on the main body. If I had the mic clip mounted on the dash, I would use the jack on the remote head.

"I'm sure there are many other options for installing a modern, light weight, and small dual-band or even HF rig, but this method proved to be satisfactory for me and upon removal, my treasured pickup truck will still be pristine with no new holes."

Photo 7. N6ZE/Rover in the Santa Monica Mts, CA during the Spring 2012, 222 MHz Sprint Contest.

Thanks, Grady! Your clever installation should inspire many "no-holes" projects, much to the relief of nervous hams and their XYLs. Readers are reminded that Grady's tips apply to his particular vehicle. Be sure all cables are properly routed away from sources of heat and moving parts as well as computer modules, and other wiring whenever possible. (Also see photos 5 and 6, K4ELO's very clean-looking installation on his pickup truck.)

Winter Mobile Operation

Most drivers know that the extremes of winter operations stress your vehicle. Pre-winter and mid-winter checkups are important; making sure your antifreeze, battery and charging system, lights, wipers, and that all-important washer fluid supply is ready for foul weather. However, have you thought about your radio gear? Nearly every radio comes with an owner's manual that also lists the radio's performance specifications, including its temperature tolerances. A transceiver that's too cold, or too hot, can do some wacky things, like transmit off the displayed frequency (perhaps even out of band), or have its display show erroneous information. If you're in an area where temperatures hit extremes, it's a good idea to let your

NOAA Weather Frequencies

Here's a handy list of NOAA Weather Radio Frequencies to program into your transceiver:162.400, 162.425, 162.450, 162.475, 162.500, 162.525.

All frequencies are in MHz; FM mode.

rig warm up a bit before transmitting. Your mid-winter check should also inspect power and coax cables, antenna connections, and grounds. Another good practice is to set your rig's time-out-timer to shut it down after a period of disuse. Some transceivers also come with a third power connection that shuts down the radio when power from the ignition switch is cut. Either of these methods can help prevent you from coming out to that most dreaded winter automotive malady—the dead battery. If your car does require a "jump start" be sure the mobile rigs are shut down, as some pretty significant voltage spikes can occur when your battery is getting a boost. A cold engine start can pull voltage down pretty low, followed by a spike when it starts and the charging system roars to life.

One more tip: Many VHF transceivers have the ability to receive the NOAA weather radio frequencies. It's a good idea to program those into some of the extra memory channels most mobile and hand-held radios seem to have (see sidebar list).

Rover?

For readers in less chilly climates, the art of VHF/UHF mountain-topping is a ticket to adventure and a lot of contacts. Grid-square and WAS award-chasers value these intrepid individuals who literally "head for the hills" to provide contacts for others. A longtime friend who often crests the hilltops is Peter Heins, N6ZE (see photo 7).

While I don't recommend an alpine adventure during the winter months, mountain-topping can be an exciting and rewarding mobile/portable pursuit, especially for those who enjoy camping and the great outdoors.

Also, if you're searching for WAS or *CQ's* coveted USA-CA All Counties Award, join the County Hunter's group usually found on 14.336 MHz. Beside the contacts, the group listens for requests for assistance from motorists across the USA, something to keep in mind during the winter travel months.

Travel safely and remember that spring isn't that far away!

73 & Happy Mobiling . .
Jeff, AA6JR

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BY TOM SMERK,* AA6TS

An Anniversary in the UK, FCC License Exams in South Africa, and Philippines Typhoon Wrap-up

I enjoy locating and reading old amateur radio magazines from the 1940s through '60s. It is amazing how much our hobby has changed in some respects and remained the same in others. Just when everyone was thinking that ham radio would "fade away," we remain strong and connected to other hams through our interests. Our first story is about a notable group in the UK that recently celebrated its 60th anniversary.

UK Emergency Group Marks 60th Anniversary

Just over sixty years ago—on November 25, 1953—the UK volunteer emergency communications group RAYNET (The Radio Amateurs' Emergency Network) was started. At the time, the UK had no such organization, and the need arose from a devastating storm that year that caused extensive damage and the loss of 307 lives. Evacuations were hindered by the lack of effective communications, and ham radio operators filled the gap.

Today, RAYNET serves as the UK's premier volunteer communications group with 2,000 licensed amateur radio operators as members (see: <http://www.raynet-uk.net/main/index.asp>). In an effort to advance emergency communications as well as amateur radio, RAYNET and the RSGB (Radio Society of Great Britain) have entered into a formal agreement to work together. RAYNET not only provides vital communications for emergency responders but also assists with public-service events.

Over the years, RAYNET has been activated numerous times, including the Lockerbie air crash in 1988, several floods, and a major oil spill.

Concerning RAYNET, Chairman Cathy Clark, G1GQJ, was quoted by the *Yorkshire Post* as saying: "... despite advances in technology, (RAYNET) is needed now more than ever. With our current unpredictable climate and the high risk of failure of modern communications networks RAYNET volunteers can make a crucial difference."

[Nevada Ham Radio.com and Southgate Amateur Radio News]

And also happening in the UK ...

West London Radio Fair at Kempton Park

The West London Radio & Electronics Fair has been held at Kempton for the last 11 years and attendance continues to grow each year. The most recent event, held in conjunction with the Radio Society of Great Britain (RSGB), took place on November 10, 2013 and was well attended by dealers, clubs, and amateurs from across the country.

The air-conditioned main hall of more than 8200 square feet attracted many of the clubs in the region to take a space along with flea-market traders and charities and included the regionally popular "bring and buy sale" where people bring items for sale and buy those that others have brought. Also popular were the many lectures and seminars and competitions, including the TOP HAM contest.

[RadioFairs.co.uk]

New VE Team to offer FCC License Exams in South Africa

The United States Federal Communications Commission (FCC) will issue an amateur radio license to anyone with a US mailing address—no residence or citizenship is required. One of the reasons for a ham living in another country to possess an FCC license is that it enables you to operate from countries with which the US has a reciprocal agreement even if your home country does not have such an agreement. It also allows you to operate within the US, even if there is no agreement in place between your country and ours.

Recently, four members of the C82DX DXpedition team who were operating in Mozambique in October, 2013 agreed to assist Chris Burger, ZS6EZ, a South African resident, in conducting an FCC testing session near Pretoria, South Africa. C82DX was an international team consisting of

*17986 Highway 94, Dulzura, CA 91917
e-mail: <aa6ts@cq-amateur-radio.com>

members from the United States, South Africa, Canada, and Japan (see <http://c82dx.com>).

Six South Africans qualified for American licenses—three for General Class and three for Amateur Extra. The newly licensed South African hams have planned to become VEs (Volunteer Examiners) in order to establish a permanent testing team in Gauteng, one of the nine provinces of South Africa.

A second exam was planned for December 2013. If you are interested in participating in a future testing session in South Africa, please contact Chris Burger, ZS6EZ, ARRL Volunteer Examiner via his QRZ.com page.

[ZS6EZ and South African Radio League (SARL)]

Philippine Typhoon Update

According to the Philippines Amateur Radio Association (PARA), when Typhoon Haiyan (locally named Yolanda)—the fiercest typhoon ever recorded on land and 3.5 times the strength of Hurricane Katrina—struck the central Philippines on November 8, 2013, killing at least four people and disabling virtually all communications services, the HERO (Ham Emergency Radio Operator) network of volunteers immediately began helping authorities with reports and messages and by tracking the storm and related outages of communications and other services. Reports estimate that more than four-million people were affected by the typhoon (see additional coverage in last month's "Public Service" column).

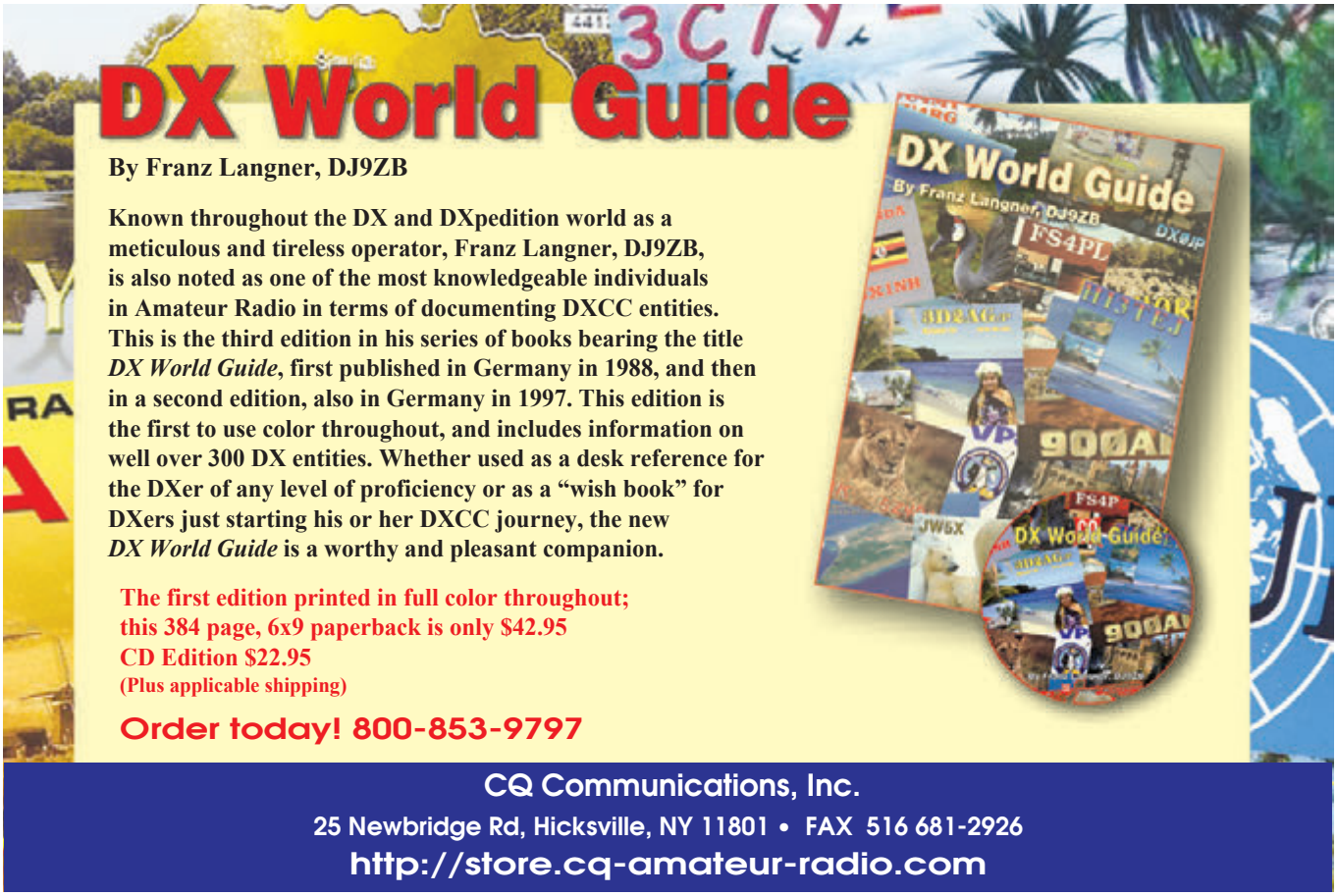
Ramon Anquilan, DU1UGZ, of PARA explained that amateur radio operations were conducted on 7.095 MHz, 144.740 MHz, and via social media. Ramon said that RADNET with Nathan Eamiguel, DU5AOK, Vilma Eamiguel, DU5VIE, and the members of their local club were operating an HF station located on the second floor of the Tacloban City Hall, powered by a generator maintained by the local government. (Tacloban is the capital of the Philippine province of Leyte and one of the places hardest hig by the storm.) Two-meter band communications were limited to simplex because there was no electricity to power their repeater.

In addition to the wind damage from the storm, flash floods and landslides caused even more misery, sending millions of people in 20 provinces to shelters, all of this occurring only a month after a 7.2-magnitude earthquake in October killed 222 people and left at least 5,000 sheltering in tents. The HERO Network also assisted in that disaster.

At least five amateur radio operations were active throughout the region, some in areas where over 90% of the structures were leveled. Many more stations receiving outgoing traffic from Tacloban and the other affected areas were scattered throughout the Philippines. Other stations were also active in receiving outbound welfare traffic, mainly to notify family members and relatives of their conditions.

DU1UGZ also reported that more than a week after being hit by the typhoon, PARA representatives attended a meeting chaired by the Commissioner of the National Telecommunications Commission, Gamaliel A. Cordoba, at which extended coverage was requested. Ramon, DU1UGZ, said, "Given the new task that NTC wants us to do, we will be needing stations that can be deployed and dismantled at a moment's notice. On the excellent offer of the ARRL, I have requested it to provide at least four HF stations and a repeater."

New HF and VHF stations were set up, and when necessary, equipment was replaced with new equipment. Lester,



Members of the C82DX team, shown here during their DXpedition to Mozambique, helped administer FCC amateur license exams in South Africa, where newly licensed Extras are starting their own VE team. (Courtesy C82DX website)

DV5PO, said that in one of the worst hit areas of Tacloban that has lost 90 percent of its buildings, the NORAD7 (Negros Oriental Radio Assistance Dumaguete) team had provided an additional HF station.

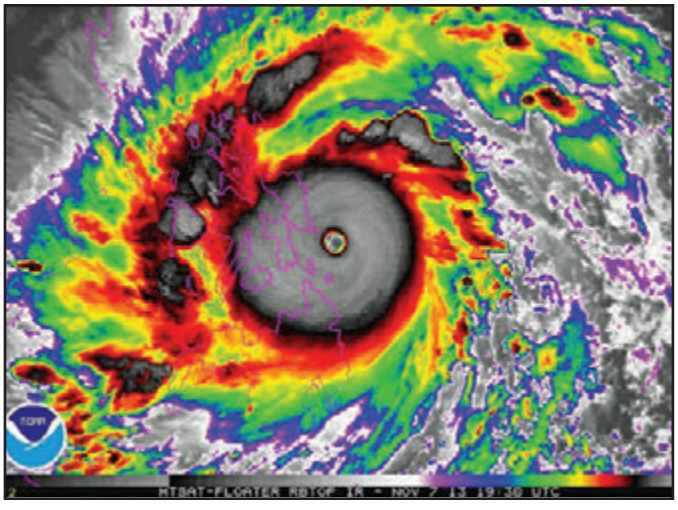
Assistance from many countries in the form of monetary donations, supplies, medical assistance, and manpower also began arriving. Earlier, the Philippine Consulate General in Guam and the United Filipino Organization said they would be starting fund drives to help victims of the super typhoon. Cell phone and Internet service were beginning to be restored as this was written, taking a lot of pressure off the ham volunteers, as victims and their families now have more options for keeping in touch and exchanging welfare reports.

Two weeks after the disaster, at the request of communication authorities in the Philippines, PARA began expanding locations and facilities and was hoping that requests for emergency communication gear, more repeaters, and even things such as more fuel for generators would be honored by the government and would provide the means for PARA's HERO network. Ramon, DU1UGZ, PARA's Vice-CEO, said the group had been working with the National Telecommunications Commission (NTC) and the National Disaster Risk Reduction and Management Council.

[Various news sources]

AA6TS comment: *People in the Philippines and other countries recently have had the unfortunate opportunity to deploy amateur radio operators to provide communications support under the worst of conditions and have had to meet extreme demands required by each situation. If any good is to come from these tragedies, it is my hope that emergency communication and disaster service volunteers from all countries are paying attention and taking notes, observing all of the unique situations that arose and noting how they were dealt with effectively so that each of us is better educated and prepared to deal with situations that might affect us at home. This author would like to extend sincere well wishes and gratitude to all of the amateur radio operators who volunteered their assistance.*

(Editor's note: As we prepared this issue for publication, we learned from ARRL that the Philippine HERO Network formally stood down on November 27 after several weeks of



Super typhoon Haiyan—one of the strongest storms ever recorded—left a path of death and destruction as it crossed the Philippines. Hams helped provide communications until normal channels were restored. (NOAA satellite photo)

Cebu Island in the Philippines (highlighted) is the home of a new RMS (Radio Message Server) system to provide digital amateur radio links to the Internet in the event of future large-scale communications outages. (CIA World Factbook map)

service, although four stations remained active in case additional needs arise.)

And in a related story . . .

Ham Radio RMS in the Philippines

In the wake of a major earthquake and a devastating typhoon, Jakob, DK3CW, reports that an RMS (Radio Message Server) has been installed on Cebu Island in the Philippines with WINMOR. Pactor (I/II and III), and other HF bands will be added later on, he says, as well as integration into the radio-only-WinLink network, to forward messages to another RMS outside the affected area if the communication infrastructure (including Internet) fails on Cebu Island.

The only other nearby WINMOR RMS is located in Australia (VK8AB), and propagation tests determined that VK8AB is not a reliable WINMOR RMS for the Philippines because of a high amount of timeouts and retries. Another RMS located in Indonesia (YBØAJZ) is only accepting PACTOR but not WINMOR for inbound sessions.

Frederick D. Medina, 4F7FDM, will be running the new system 24/7 using WINMOR 500 on 7036.5-kHz center frequency and VHF 9600-baud packet radio.

[DK3CW via Southgate Amateur Radio News]

Indian fishermen Want Free Ham Radio and GPS Gear

If you haven't heard about this yet, I want to share a story that I found both imaginative and perhaps disturbing at the same time—AA6TS:

According to India's *Deccan Chronicle*, there is a group of fisherman in that country that is upset about the lack of affordable and reliable communication equipment that they can use while deep-sea fishing, especially when they face severe storms. Currently, their only way to communicate location or safety-related information to their families is with mobile telephones that they claim are often unable to obtain a signal and have limited back-up power. Many are using portable FM broadcast radios to receive weather information.

Around 70,000 fishermen from several villages along India's coast have been demanding the state government provide them with free amateur radio gear and GPS equipment (as a subsidy) because they cannot afford to purchase this equipment on their own. They say that if they are equipped with amateur radio and GPS, it will help to get weather updates, receive directions, and find their location at sea to avoid impending danger.

The fishermen are backed by the president of the District Fisherman Welfare Association that was quoted in the article as saying that when the communication network fails during the storm, fishermen find no source of information unless they are equipped with ham radio and GPS to help them reach shore safely. The fishermen are awaiting a response from the government to their demand.

What caught my attention is the fact that the article did not mention any discussion of *licensing*, and I have a hard time believing that 70,000 fishermen are newly licensed but have not obtained their equipment yet. I will continue to watch this development with interest because I feel that the IARU would never condone the requested amateur radio equipment to be issued to unlicensed operators. I would delight in reporting in a future issue about the fact that India has added 70,000 new licensed amateur radio operators, not to mention the other family members at home! *(Another potential problem here is that the fishermen would improperly use ham gear for business-related communications.—ed.)*

Fishermen and boaters from all countries have discovered the benefits of obtaining an amateur radio license and having ham gear on-board for personal and emergency communications, but the fascinating part of this report is the huge number of potential new hams that are unified in their efforts to obtain amateur radio.

[Deccan Chronicle]

Until next month, please continue to share amateur radio news and happenings from your country. Send your stories and photos to <aa6ts@cq-amateur-radio.com>.

73, Tom, AA6TS

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BY JOE LYNCH,* N6CL

Teenager Leads ISS Contact

School contacts with the International Space Station have become so routine that most of the time they are no longer covered in places such as this column. Therefore, it is news when something special happens during one of those contacts.

Something special did happen during the November 13, 2013 ISS contact: Sixteen-year-old Rebecca Rubsamen, KJ6TWM, was the lead engineer for the contact between students at Rancho Romero Elementary School and astronaut Mike Hopkins, KF5LJG. Rubsamen of Alamo, California, is a sophomore at Bentley School in Lafayette and built her own Elecraft™ K3, which included 2 meters. In her backyard she also built the two antennas used in the contact.

Rebecca's interest in amateur radio got started by her father, Reid Rubsamen, N6APC, two years before the contact. Not to be content with ordinary QSOs, she learned about the ISS and satellite communications so that she would be able to return to her former elementary school and lead this ISS QSO.

As we know, it is a tough process to be selected for an ISS contact. NASA and the Amateur Radio on the International Space Station (ARISS) arrange for about 50 contacts per year. Undaunted by her age, she applied for and received permission to be one of the youngest lead operators for an ISS contact.

According to Ashle Harris, a NASA spokes person, last year there were around 68 permissions granted. While most of them were international contacts, 20 were with stations in the U.S., of which three (including this one) were with California stations.

While Tim Bosma, W6MU, was the ARISS-assigned mentor for the contact, Rebecca ran everything from arranging the station's setup to making sure that the students were in queue so they could ask their questions. Tim was quite impressed with her professionalism, adding that he had not seen anything quite like it in his approximately thirty years of being a mentor. During the nearly 8-minute QSO most of the 15 students were able to ask Mike their questions. The audio of the contact can be found at: <http://www.youtube.com/watch?v=dDv0m-7C4Oo>.

Rebecca maintains a website (http://www.rebeccarubsamen.com/photos.html) in which she has posted a number of photos pertaining to the run-up to the launch. Among the photos is one with astronaut James Van Hoften and her on Science Night at her school. In that photo she displays her K3 and the 12-volt power supply that she used to power it. Local media coverage was supplied by the Contra Costa Times. Here is a link to its story: <http://www.contracostatimes.com/news/ci_24517766/headline>.

While it may be a tad bit too early for Rebecca to think about her career aspirations, it is clear that she already knows how to achieve what she sets her mind to do. Look for her to make more headlines in the not too distant future. Hopefully for our hobby, she could be making those headlines in connection with amateur radio.

VHF Plus Calendar

February 6	First quarter Moon
February 12	Moon apogee
February 14	Full Moon
February 22	Last quarter Moon
February 27	Moon perigee

Hector Martinez, CO6CBF, now also KF5YXV

This past November Hector Martinez, CO6CBF, earned his U.S. Technician class amateur radio license. He is probably the only Cuban national to also hold a U.S. amateur radio license. Here, from Southgate Amateur Radio News service is how he was able to become a licensed U.S. amateur radio operator:

The idea was first conceived during the 30th AMSAT Space Symposium in Orlando 2012 by Clayton Coleman, W5PFG, after hearing Hector Martinez, CO6CBF's desire to operate with his own callsign during his visit to the United States. Because there is no reciprocal agreement between the U.S. and Cuba, he was not able to do so.

A workaround was for Hector to earn a U.S. amateur radio license. At that time he was able to identify three Volunteer Examiners. However, they lacked the materials to test Hector at that time.

Fast forward one year to the 31st AMSAT Space Symposium in Houston.

A few weeks before the Symposium, Clayton contacted the AMSAT-BB, asking for AMSAT members who were ARRL or W5YI VEs to contact him in order to see if we might be able to set up a testing session for Hector. Clayton received many responses and four ARRL VEs stepped up

Shirley, N8LX, and Lynn, N8LXK, Roberts at the Handiham booth during the 2011 Hamvention™. See note at end of this column. (Photo courtesy Handiham)

e-mail: <n6cl@sbcglobal.net>

to assist the effort. Douglas Phelps, K9DLP, took the lead as Team Liaison and brought testing and filing materials with him. Others joining the team included EMike McCardel, KC8YLD, Patrick Stoddard, WD9EWK, and Jerry Buxton NØJY.

Late Thursday evening, rather serendipitously, three of the VEs happened to find themselves in the same place at the same time. Shortly thereafter Clayton, W5PFG, Patrick, WD9EWK, and Hector also happened by. The team immediately went into planning mode to set up a time and place to conduct the test, preferably early Friday. Then came the suggestion, why wait? Everyone is here, so why can't we do it now? When the team asked Hector if now would be good, his immediate response was, "Yes, yes, now!"

The team went to their rooms to collect their credentials and materials and then convened 15 minutes later in a quiet, well lit area of the hotel lobby and presented Hector with a copy of the Technician Exam. Hector took it and passed the exam.

The next morning the exam materials and application were forwarded on to the ARRL VEC by means of US Postal Priority mail, which assured delivery by Monday noon, in hope that his license would be granted before his return on Wednesday. At shortly before 3 pm CST Monday Hector's license cleared the FCC ULS database and he officially was granted callsign KF5YXV. We can't laud Maria Somma, AB1FM, and the folks at the ARRL enough for their efficiency.

Minutes after all became aware that Hector had been assigned his callsign, he began working the birds. Clayton reports, "We had a blast on Monday evening through Wednesday working every satellite pass available to us. He was as exciting to watch as a kid getting his license for the first time."

Hector adds, "Thanks very much for all your help and the great time! It was very fun to operate for first time on AO-7 mode B."

Congratulations to Hector and all who participated in this expression of International Goodwill.

Thanks go to Clayton Coleman, W5PFG, EMike McCardel, KC8YLD, and AMSAT News Service for the above information.

FUNCube-1 Satellite Launched

During the middle of November, 37 satellites carrying amateur radio payloads were launched. Among them was the much anticipated FUNCube-1 satellite, now designated AO-73. Here is the AMSAT-UK FUNCube website: <http://funcube.org.uk/>.

VK Waits for Full Return of the 6-meter Band

The following, from Jim Linton, VK3PC, by way of Southgate Amateur Radio News Service:

With only digital television being available in Australia from 10 December, and the loss of the last TV transmitter Channel 0, it does not automatically mean the unfettered return of 50–52 MHz to the amateur radio service.

Restrictions listed in the Licence Conditions Determination (LCD) on operation of amateur stations in Australia's eastern states on the lower end of 6 metres remain effectively in force until the document is changed by the Australia Communications and Media Authority (ACMA).

Channel 0 came on air in Australia in the 1960s, which meant radio amateurs lost the bottom 2 MHz of 6 metres.

Australia expanded its TV analogue allocation because of the need for more channels and so was born 45–50 MHz or Channel 0, initially for commercial telecasters, then the Special Broadcasting Service, and for translators.

The Melbourne and Brisbane telecasters left Channel 0 and became Channel 10, while the gap was filled by SBS, with it eventually settling on UHF Channel 28.

The Wireless Institute of Australia (WIA) has long argued that the 50–54 MHz band should be retained for, and allocated exclusively to, the amateur service, once the existing Channel 0 transmitters are removed.

The 50-MHz or so-called "magic band" is important to radio amateurs because it provides unique opportunities for propagation experiments.

The WIA has again written to the ACMA seeking the unrestricted return of the band. The latest letter on the subject last month is yet to be answered.

The ACMA reminded all in July this past year of their LCD obligations in relation to the 6-meter restrictions.

On previous occasions the limited resources of the ACMA have meant the change waits until the Australian Radio Frequency Spectrum Plan is amended, and that usually occurs in January.

The WIA has made a number of submissions to the ACMA on the subject over the years seeking a harmonization in Australia with the practice overseas.

The WIA is also to amend the Regulations licence assessment questions to reflect any change, while at the same time it will beef up its electromagnetic radiation emphasis in the papers from 2014.

Radio Amateurs Say Hi To Juno

The following is from AMSAT-UK (http://amsat-uk.org/2013/10/09/radio-hams-say-hi-to-juno/):

Radio amateurs around the world took part in an experiment with NASA's Juno spacecraft as it did a flyby of Earth. NASA's Juno spacecraft flew past Earth on Wednesday, October 9, 2013 to receive a gravity assist from our planet, putting it on course for Jupiter.

To celebrate this event, the Juno mission invited amateur radio operators around the world to say "HI" to Juno in a coordinated Morse Code message that would be detected by Juno's radio and plasma wave experiment, called Waves.

Radio amateurs transmitted Morse (CW) signals on a range of frequencies between 28.001 and 28.450 MHz. To give a random spread the precise frequency used depended on the last character of each station's call sign. The natural signals the team expect to measure at Jupiter will consist of a large number of discrete tones, so spreading the signals out in this manner was a good approximation to the signals Juno is expected to detect. But at Jupiter, they don't expect to be able to decode CW in the telemetry!

The Waves instrument is sensitive to radio signals in all amateur bands below 40 MHz. However, experience with the University of Iowa instruments on the Galileo and Cassini Earth flybys showed significant shielding by the ionosphere at lower frequencies, so the 28 MHz band was chosen for the experiment.

Juno's antenna consists of a pair of tapered 2.8 meter long titanium tubes, deployed from the bottom deck of the spacecraft under the +X solar array and magnetometer boom. A high impedance radiation resistant preamp sits at the base of the antenna and buffers the signals from 50 Hz

to 45 MHz. The elements are deployed with an opening angle of about 120 degrees. 28 MHz is above the resonant frequency of the antenna and NEC analysis indicates a lobe generally along the spin axis of the spacecraft. This will be good for detection on the inbound part of closest approach to Earth.

The Waves instrument uses four receivers to cover the frequency range of 50 Hz to 41 MHz. Signals up to 3 MHz are bandpass filtered, sampled by A/D converters and FFT processed into spectra using a custom FFT processor developed by The University of Iowa under a grant from the Iowa Space Grant Consortium.

Among those taking part were students at Virginia Tech using their club station K4KDJ. Here is a video of that effort: <http://www.youtube.com/watch?v=OaCRnSQf9sE>.

More on Juno

The following is from Southgate Amateur Radio News:

When NASA's Juno spacecraft flew past Earth on October 9, 2013, it received a boost in speed of more than 7.3 kilometer per second, which set it on course for a July 4, 2016 rendezvous with Jupiter, the largest planet in our solar system

During the flyby, Juno's Waves instrument, which is tasked with measuring radio and plasma waves in Jupiter's magnetosphere, recorded amateur radio signals. This was part of a public outreach effort involving ham radio operators from around the world. They were invited to say "HI" to Juno by coordinating radio transmissions that carried the same Morse-coded message. Operators from every continent, including Antarctica, participated. The results can be seen in this video clip at <http://photojournal.jpl.nasa.gov/archive/PIA17744.mov>.

Watch a four-minute video depicting the efforts of a few of the many amateur radio operators who participated in the event here: <http://www.youtube.com/watch?v=_yqHy_MpNiQ>.

One of Juno's sensors, a special kind of camera optimized to track faint stars, also had a unique view of the Earth-moon system. The result was an intriguing, low-resolution glimpse of what our world would look like to a visitor from afar.

"If Captain Kirk of the USS Enterprise said, 'Take us home, Scotty,' this is what the crew would see," said Scott Bolton, Juno principal investigator at the Southwest Research Institute, San Antonio. "In the movie, you ride aboard Juno as it approaches Earth and then soars off into the blackness of space. No previous view of our world has ever captured the heavenly waltz of Earth and Moon."

Watch the Juno Earth flyby movie with a music accompaniment by Vangelis here: <http://www.youtube.com/watch?v=CzBlSXgzql>.

The cameras that took the images for the movie are located near the pointed tip of one of the spacecraft's three solar-array arms. They are part of Juno's Magnetic Field Investigation (MAG) and are normally used to determine the orientation of the magnetic sensors. These cameras look away from the sunlit side of the solar array, so as the spacecraft approached, the system's four cameras pointed toward Earth. Earth and the Moon came into view when Juno was about 600,000 miles (966,000 kilometers) away—about three times the Earth-Moon separation.

During the flyby, timing was everything. Juno was traveling about twice as fast as a typical satellite, and the spacecraft itself was spinning at 2 rpm. To assemble a movie that wouldn't make viewers dizzy, the star tracker had to capture a frame each time the camera was facing Earth at exactly the right instant. The frames were sent to Earth, where they were processed into video format.

"Everything we humans are and everything we do is represented in that view," said the star tracker's designer, John Jørgensen of the Danish Technical University, near Copenhagen.

"With the Earth flyby completed, Juno is now on course for arrival at Jupiter on July 4, 2016," said Rick Nybakken, Juno project manager at NASA's Jet Propulsion Laboratory in Pasadena, California.

The Juno spacecraft was launched from Kennedy Space Center in Florida on August 5, 2011. Juno's launch vehicle was capable of giving the spacecraft only enough energy to reach the asteroid belt, at which point the sun's gravity pulled it back toward the inner solar system. Mission planners designed the swing by Earth as a gravity assist to increase the spacecraft's speed relative to the sun, so that it could reach Jupiter. (The spacecraft's speed relative to Earth before and after the flyby is unchanged.)

After Juno arrives and enters into orbit around Jupiter in 2016, the spacecraft will circle the planet 33 times, from pole to pole, and use its collection of science instruments to probe beneath the gas giant's obscuring cloud cover. Scientists will learn about Jupiter's origins, internal structure, atmosphere and magnetosphere.

Source:<http://www.jpl.nasa.gov/news/news.php?release=2013-360>.

And Finally . . .

Some of you who have seen my wife, Carol, W6CL, and me at the Dayton Hamvention™ have also seen us with Shirley, N8LX, and Lynn, N8LXK, Roberts. Shirley was Hamvention's Ham of the Year in 2011. Now Shirley is battling breast cancer. I ask you to please keep her in your thoughts and prayers. Also, please send her an e-mail of encouragement to: <n8lx@earthlink.net>.

This is all the news I have room for in this issue. Thanks to all . . .

73 de Joe, N6CL

what's new

Firmware Upgrade for ICOM IC-7800 Transceiver

ICOM America has released a new firmware upgrade for its popular IC-7800 HF/6-meter transceiver. Features include:

- Spectrum Waterfall Display
- PC Mouse Operation via USB
- Audio Scope Function
- Direct Remote Control Operation via the Internet
- Digital Voice Recorder

The firmware upgrade may be downloaded from the IC-7800 product page on the ICOM website at <http://bit.ly/JIfKEe>.

Note: *"What's New" is not a product review and does not constitute a product endorsement by CQ. Information is primarily provided by manufacturers/vendors and has not necessarily been independently verified. The purpose of "What's New" is to inform readers about new products in the marketplace. We encourage you to do additional research on products of interest to you.*

BY TED MELINOSKY,* K1BV

QRP Awards, plus a Very Short-Term Award

It's usually difficult to be able to feature very short-term awards in this column because the column is written well in advance advance of the time you receive your copy of CQ. However, first is an exception. The Punxsutawney Area Amateur Radio Club has published its rules to offer its annual "Ground Hog Day" certificate (it's free!).

For our overseas readers, the "Ground Hog" is also known as a woodchuck, a rodent belonging to the group of large ground squirrels known as marmots. The woodchuck is a lowland creature. It is widely found in North America.

Ground Hog Day is celebrated on February 2. According to folklore, if it is cloudy when a groundhog emerges from its burrow on this day, then spring will come early; if it is sunny, the groundhog will supposedly see its shadow and retreat back into its burrow and the winter weather will continue for six more weeks. (Thanks Wikipedia)

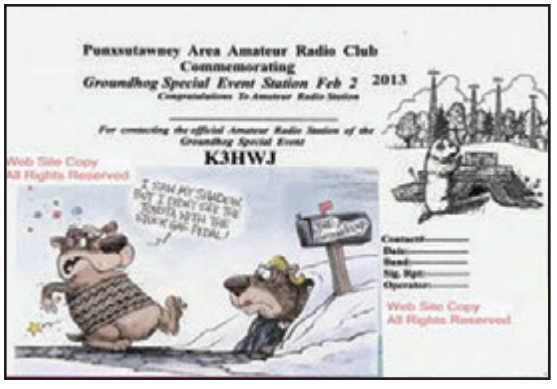
USA Ground Hog Day Special Event Certificate

Make just one contact with the Punxsutawney Radio Club station K3HWJ on Saturday, February 2, 2014 from about 9:00 a.m. until at least 3:00 p.m. or later depending upon band conditions. Look for activity on the following frequencies:

40 meters on or about 7183 ±about 5 kHz
20 meters on or about 14330 ±about 5 kHz
80 meter backup on or about 3845

Send your QSL card and a self-addressed, stamped envelope to: Punxsutawney Area Amateur Radio Club, Groundhog Contact No. ___, P.O. Box 3, Punxsutawney, PA 15767. Note: Put your contact number in the "___" above. This will help the club to process the certificate.

*12 Wells Woods Rd., Columbia, CT 06237
e-mail: <k1bv@cq-amateur-radio.com>



The Punxsutawney Area Amateur Radio Club has published its rules to offer its annual short-term "Ground Hog Day" certificate

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The total number of counties for credit for the United States of America Counties Award is 3077. The basic award fee for subscribers is $6.00. For nonsubscribers it is $12.00. To qualify for the special subscriber rate, please send a recent CQ mailing label with your application. Initial application may be submitted in the USA-CA Record Book, which may be obtained from CQ Magazine, 25 Newbridge Road, Hicksville, NY 11801 USA for $2.50, or by a PC-printed computer listing which is in alphabetical order by state and county within the state. To be eligible for the USA-CA Award, applicants must comply with the rules of the program as set forth in the revised USA-CA Rules and Program dated June 1, 2000. A complete copy of the rules may be obtained by sending an SASE to Ted Melinosky, K1BV, 12 Wells Woods Road, Columbia, CT 06237 USA. DX stations must include extra postage for airmail reply.

Internet: <http://www.punxyclub.com/special-events/ground-hog-day>

QRP Awards

Have you ever tried QRP? Most modern rigs will let you crank the power down to 5 watts or less for tuning purposes. Here are some QRP awards from the major and minor clubs that specialize in this area of ham radio.

The first is from the QRP ARCI International (QRPARCI) , probably the largest club of its type in the USA. It offers a dozen awards. Follow the link below in the rules to view all offered.

USA QRP ARC Series – Worked All States Award. QRP is defined by the QRP ARCI as using no more than 5 watts output on CW and 10 watts PEP output on SSB. This award is available to any amateur operator. Application forms may be found on the club's website: <http://www.qrparci.org/arciawds.html>. The fee for all awards or endorsement certificates for W/K amateurs is $US4, or for non-W/K $US5 or 10 IRCs. In addition, e-QSLs are now accepted. There are endorsements for band and mode, etc. Apply to:

The QRP ARCI Series Worked All States Award.

On the Cover

Dave Clausen, W2VV, is a co-founder of the NYC Resistor in Brooklyn, New York—the first "makerspace" in the United States—and trustee of its club ham station, N2YCR, where he is operating in the photo.

A ham since 2008, Dave's main interests are QRP (appropriate for our QRP Special!), portable operating, and activating summits for SOTA, the Summits on the Air program. He also enjoys building kits.

Dave's interest in ham radio came out of his involvement in NYC Resistor, which he helped to found in 2007. "I had no particular interest in radio," he recalls, noting that his main interest at the time was working with microcontrollers. "But one guy in the club had his license; I got mine, and so did others." Dave notes that the maker group is not a ham radio club as such, although there is a ham club that's part of it (N2YCR), and that there's "lots of overlap."

Dave says he sees more interest and excitement about ham radio among younger people today than five to ten years ago, and thinks one way to recruit more young people is through broader-based groups, such as a maker group, that include ham radio as part of their activities.

For readers who are not familiar with makerspaces or hackerspaces, Dave describes NYC Resistor as part clubhouse, part social group, part workspace with shared tools, and part business incubator. Several businesses have gotten their start there, he explains, most notably MakerBot, of which fellow co-founder Bre Pettis, W2BRE, is CEO. The group sponsors classes and holds weekly craft nights, which are open to the public as well as to members.

A software engineer for Google, Dave also has been the cameraman for many of the popular ham radio videos produced by his wife and fellow NYC Resistor co-founder Diana Eng, KC2UHB. And yes, they met at the club!

(Cover photo by Larry Mulvehill, WB2ZPI)

Just one QSO, of 100 kilometers or more, using 1 watt or less, qualifies you for the QRP$_p$ Award.

QRP ARCI Awards Chairman, QRP ARCI Awards Manager, Julio A Jimenez ,3341 Eastgate Street, Eastover, NC 28312, USA.

E-mail: <awards@qrparci.org>, Internet:<http://www.qrparci.org/content/blogsection/4/116/>

QRP All States Award. This award is issued to any amateur for confirming QSOs with stations in 20 or more of the 50 states of the USA while running QRP. A satellite endorsement is available. Endorsement certificates are issued at 30, 40, and 50 states confirmed. A satellite endorsement is available.

USA QRPp Low Power Award. Successfully using extreme low power to make contacts over long distances is always a fascinating subject. A long time ago, when I was newly licensed, I used just my VFO to call a big shot DXer a few towns away, and the antenna was a mismatched long wire. Yes, he heard me, sent QRZ? and his callsign at high power. I was convinced that QRP could really work.

This award was developed in order to encourage more low-power amateur radio operations.

1. Make one QSO using power of less than one watt. The output power during the period of qualifying communications was accurately measured to be less than one watt (QRPp).

2. The distance between the applicant's transmitting antenna and the receiving station was over 63.3 miles (100 kilometers).

3. No artificial means of active relay was used to complete the communications (i.e., repeaters, satellite transponders, digi-peaters, land-lines, etc.).

4. Send a signed statement to the custodian affirming that the transmitter power was less than one watt, the distance was greater than 100 kilometers, and no artificial means of active relay was utilized. Provide a photocopy of the station log, or photocopy of the confirmation QSL card (electronic QSL cards are not acceptable), clearly showing the date, time, mode, and frequency on which the qualifying communication took place. If desired, provide information concerning any endosements (such as longer distances, even lower power levels, WAC, WAS, WAZ, solar power, etc.) that you may want listed on your certificate.

5. Mail fee of $4.00 ($5.00 for foreign), to cover postage, paper, and printing along with the application materials shown above to: Roger J. Wendell, WBØJNR, QRPp Certificate Manager, P.O. Box 17174, Golden, CO 80402-6019 USA. Internet: <http://rogerwendell.com/qrpaward.html>

Italy's I-QRP Award. I like this one from Italy. Both sides of the QSO do not have to be made using QRP. It's somewhat hard to contact stations that both must use QRP to make and complete a contact. Here, if you are the QRP station and contact the Italian station who is using non-QRP, you still get credit, if only 1 point. And there are many, many Italian stations on the air, so your chances of being heard, under good conditions, with QRP are pretty good.

Earn a minimum of 50 points by contacting at least 25 Italian QRP stations. If both stations use QRP, each QSO = 2 points; if only you use QRP, the contact is worth 1 point. All bands may be used, including WARC. Each station may only be contacted one time. Send GCR list and fee of 7 Euros, $US7, or 7 IRCs to: Marcello Surace, IK7HIN, via Dante 239, I-70122 Bari, Italy.

E-mail: <marcello.surace@tiscali.it>; List of I-QRP members: <http://www.arimontebelluna.it/i_qrp/contest.htm>

We are always interested in learning of new awards for this column. Please contact me with any details; a URL on the internet would be fine.

73, Ted, K1BV

Italy's I-QRP Award.

Paul, N7JP, USA-CA All Counties #1238

Paul H. Nelson, N7JPF
USA-CA All Counties #1238
October 24, 2013

Isn't ham radio a fantastic hobby? There are so many different types of transmission modes and interesting areas a person can focus on. County Hunting is an area I started to work on many years ago and have never regretted the trip.

I was introduced to ham radio by some of my father's friends. My dad served in the Army Signal Corps during WW II, stationed in Alaska working in the Alaska Communication System. He wore headphones many hours a day for three to four years, hearing incoming code and typing it on the "mill." When he came home from the war, he was not interested in pursuing ham radio like many of his other signal corps buddies. My first radio recollection was visiting the home of one of his wartime friends and seeing his station. "Mack" Mackie, W7IJY(SK), was the first operator I ever knew, and I will never forget the warm glow of his station and the places he could talk to. Three decades later, when I got my ticket, I was thrilled to be able to have a QSO with him.

In the mid-1980s, married and with a family, I met my soon-to-be Elmer, Tom, WA7TC, who got me going on the code and license, and then into homebrewing antennas. With Tom, my ham skills were further practiced in ARES/RACES, SAR, and American Red Cross communications in Grays Harbor County, Washington.

Thumbing through a copy of CQ magazine one day, I became interested in the several awards that were offered by confirming contacts. I chose to work on the CQ USA-CA Counties Award, and spent the next 22 years toward its completion. There were many years of interruption because of moves, and a job that required many hours away from home, not to mention the obligation of raising a young family. Chasing counties is a hobby that can be put aside and come back to easily when you have time. I love the county hunting nets because you don't have to sit there for hours. It only takes seconds to get a needed county contact, and then you can go back to whatever you were doing.

Checking my logs, I see my first contact on the 14.336 County Hunter net was with Milt, KYØE, mobile in Apache, AZ, on May 19, 1991. My last contact, for the "whole-ball-of-wax" was with Bob, N8KIE, on October 10, 2013 from Towns, GA. Bob went out of his way in his recent trip to get my last four counties, two in KY and two in GA. Since there are 3,077 counties, there were a lot of other fantastic people between those two contacts.

Special thanks go to Bob, N8KIE, obviously. Also to Tom, WA7TC, for being such a great Elmer, and to Terry, WQ7A, and Bob, K7TM, for their help and inspiration. Several people helped with completing states in the last few months; NA8W, NØKV/NØDXE, WG9A, N8HAM, KM9X, and K5GE. I am sure there are more I have missed, and I am thankful for you as well. There are several CW operators who were very patient and forgiving of my fist, as I started to make the switch a few years ago, seeing that the path to the finish would come a lot faster on CW. A few of those great CW operators deserving mention are WØGXQ, N4CD, W7FEN, K1TKL, W4SIG, KC7YE, and K7TM. Also, special thanks to my wife, Darlene, who has listened to so much QSB as we drive, and the seemingly endless contacts as I give out a county. She has more patience than any ham operator I have ever worked.

I have met many mobile operators at the past MARAC conventions, and the friendships that have been made are wonderful. I want to thank the numerous county hunters for all the counties and the fantastic job of the net controls and relays. They are first-class operators, and they have made this possible. *. . .UR 599 EWA TU. —Paul, N7JP*

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BY WAYNE MILLS,* N7NG

A Look Back at 2013 and a Report from South Sudan

First off, I must apologize profusely for the mix-up last month. I finished the January column while spending more than 12 hours at John F. Kennedy International Airport in New York waiting to fly to Europe to join the Z81X DXpediton to South Sudan. Somehow, while traveling I failed to pass the column to CQ Headquarters. It seems I could have hand-delivered it! Once out of the country, access to the secure areas of my ISP were blocked. My ISP people said that it was normal, but I wonder...

Thus, with *this* column, I start the year 2014. This month I will include some of January's material, which was a review and update some of the topics that I talked about in my 2013 DX columns starting with April 2013. I will conclude with some thoughts—and photos—from the Z81X expedition.

April – Introduction

My April column contained a little bit about me and a little bit about my philosophy of DXing. I mentioned DXing skills and the lack thereof. What I didn't mention was the fun that I had in the early days and since. When I was much younger, talking to others in distant locations was very intriguing, and I always said that I someday would like to visit some of the people whom I meet on the radio. That "dream" has come true for me. Your prior DX column editor, Carl Smith, N4AA, often said that no matter what your focus, having fun is, in fact, what this hobby is all about. What we older DXers might regret is that newer hams our age don't have the same opportunity to appreciate such communications. However, that's life, and we need to cope with it. The young folks have their own joys, and we elders need to understand that.

May – DXing Tips, Thoughts on Kosovo

May brought some DXpeditioning tips and more thoughts on Kosovo. Recently, Rick, K6VVA, has made it a career to study pile-up behavior, and he has identified some interesting traits. I believe the most useful is that many DXers don't *really* know what's happening in a pile-up—whether the DX operator is speeding along at 39.9 words per minute or on SSB near the noise level. Communicating with the pile-up where, when, and how to call is imperative. To accomplish this, slower CW and clarity and repetition on SSB are vital. Languages and accents must be accounted for.

On CW, I don't mean 15 words per minute, but a comfortable 26–30 wpm is a good speed, slower perhaps if conditions are more difficult. Sending "UP" after every QSO is helpful, and sending your callsign at least every three QSOs is mandatory. Even using these techniques, it is impossible for everyone to know exactly what is going on if the operator isn't consistent.

On phone, it is equally important, but somewhat easier to communicate effectively with the pile-up. Keeping in mind that not all DXers in the world speak English as a first language is essential. In the end, we need to be more understanding of what is going on in a pile-up, especially the so-called "cops" who wish to keep the DX frequency under control, but usually are more QRM than those calling on the wrong frequency. If the DX op fails to inform the callers, chaos will ensue.

I also discussed the status of Kosovo. This new country is recognized by CQ and the German Amateur Radio Club (DARC). It is still not recognized by the ARRL and its DXCC program. In July, however, the ARRL Board tasked the DX Advisory Committee with reviewing *all* DXCC rules. Two of the important issues within that task are to consider country status criteria and remote-control operation. We can only presume that additional discussion will have taken place at the Board of Directors' January meeting.

And, again, I have gotten carried away until…

June – Too Much DX? And a Look at DXpedition Funding

For the June column, I lamented having to work the Pacific DXpeditions during the early hours of the morning. Four a.m. and I have never gotten along too well, and now in early November (when this column was originally written), I am again having to get used to these dark, early hours in order to scoop up whatever DX might be available in the Pacific on Topband. Ah well, whatever it takes.

The main topic in June, however, was DXpedition funding. Don, K1DG, had presented a paper describing where in the world DXpedition funds come from, and I counseled consideration of local customs and traditions regarding donations and charity.

In that column, there was an announcement of a large NCDXF grant made to the Amsterdam Island DXpedition due on the air last month. Reasonably large expenditures for such difficult destinations may be appropriate. Yet it may be that such large contributions only perpetuate the idea that the super-large-scale expeditions are feasible. The infrequency of such operations seems to beg for the largest possible range of band and modes, and the largest possible number of contacts. However, when the total cost approaches $8 to $10 per QSO, some rethinking might be in order.

*P.O. Box 1945, Jackson, WY 83001-1945
e-mail: <n7ng@cq-amateur-radio.com>

The WPX Program

CW

3465	WB9G	3479	WF2S
3469	G4FKA	3480	AF5DM
3471	R5QA	3481	AI6II
3472	F5UL	3482	WØHI
3474	WU1V	3483	NZ3D
3475	R3LC	3484	JP1EWY
3476	PY3IT	3485	KN7D
3477	YB7XO	3487	9H1SP
3478	K6CTA		

SSB

3386	IZØFUW	3417	SQ8OQE
3400	G4FKA	3418	KB3SGY
3406	KI7DG	3419	IKØOPS
3407	JR3UIC	3420	W3GLL
3409	BH4SCF	3421	AC7JM
3410	N6MM	3422	W7YED
3411	AG6OX	3423	JA1SCE
3412	N6ETO	3424	AB3TM
3413	R3LC	3425	KA8YYZ
3414	EA5FCW	3426	KG2U
3415	K1BZ	3427	WØBM
3416	XQ7UP	3428	9H1SP

Mixed

2600	R5DT	2649	AI6II
2604	IZØFUW	2650	SP5RE
2611	WB9G	2651	AA4OC
2622	DH3SR	2652	PA3HFX
2631	R5QA	2653	AB3TM
2623	I5OYY	2654	HA5NB
2639	F5UL	2655	NTØF
2640	R3LC	2656	WØHI
2641	UW1HM	2657	DD8BA
2642	K1BZ	2658	JP1EWY
2643	N3FHW	2659	KN7D
2644	SP5XOV	2660	KC9UNL
2645	N8WXQ	2662	IZ8GUQ
2646	W3GLL	2663	WØBM
2647	LZ4SU	2664	K4SBZ
2648	W8RLS	2665	9H1S

Digital

219	KØWIU	270	BX2ADE
244	JE1HRC	272	PA3HFX
267	IZØFUW	271	SP5XOV
268	PY3IT	273	KC9UNL
269	KB2GCG	274	K4SBZ

CW: 350 JE1HRC, WF2S, WB9G. 400 YB7XO, KC8UR, K6CTA, WØHI. 450 KN7D, R5QA, JP1EWY, K4MLD. 500 JR3UIC. 550 G4FKA, R3LC. 600 AB1OC, F5UL, K3CWF. 700 N4HB. 1100 9H1SP. 1200 V51YJ

SSB: 400 IKØOPS, WØBM. 450 K1BZ, IZ2BVN, IZ8GUQ. 500 F5VHQ. 550 N6ETO, G4FKA. 600 W3GLL, YBØDJ, AA4OC, N4FNB. 650 IZØFUW. 700 N4HB. 750 TA1HZ, NTØF. 850 XQ7UP. 950 KA8YYZ. 1500 AB1OC. 1550 R3LC. 1650 9H1SP

Mixed: 450 KK4FYV, IZ2BVN, WØHI, IZ8GUQ. 500 K4MLD. 550 AE4WG, R5QA, JP1EWY, PA3HFX. 600 PY3IT, JE1HRC, W3GLL, AA4OC, N4FNB, WØBM. 700 K1BZ, F5UL. 750 N8WXQ, KN7D. 800 TA1HZ. 850 IZØFUW. 900 N7ZO. 1000 JR3UIC, HA5NB. 1050 G4FKA, R5DT. 1100 K1DX. 1150 N4HB, AJ4FM. 1200 I5OYY. 1250 XQ7UP. 1300 K3CWF. 1600 R3LC, JG3FEA. 1650 WF2S. 1900 AB1OC, 9H1SP. 3550 K9UQN. 4000 K1BV. 5200 HA5DA.

Digital: 350 KK4FYV. 400 DD8BA, PA3HFX. 500 KØWIU. 550 WA5VGI. 650 JR3UIC. 850 AB1OC. 1150 WF2S.

160 Meters: K1BZ, R3LC, N4HB, WB9G, 9H1SP
80 Meters: R3LC, 9H1SP
40 Meters: R3LC, N4HB, TA1HZ, JG3FEA, 9H1SP
30 Meters: V51YJ, JG3FEA, 9H1SP
20 Meters: KK4FYV, R3LC, JR3UIC, IZØFUW, YBØDJ, N4FNB, KN7D, JG3FEA, WØBM, IZ8GUQ, 9H1SP
17 Meters: JG3FEA, 9H1SP
15 Meters: N6ETO, R3LC, JG3FEA, 9H1SP
12 Meters: JG3FEA, 9H1SP
10 Meters: G4FKA, R3LC, JG3FEA, 9H1SP
6 Meters: 9H1SP

Africa: AB1OC, 9H1SP
Asia: JE1HRC, R3LC, BD2AXE, IZØFUW, KN7D, JP1EWY, JG3FEA, 9H1SP
Europe: N6ETO, DH3SR, G4FKA, K1BZ, KK4FYV, R3LC, N3FHW, SP5XOV, F5VHQ, IZØFUW, WF2S, YBØDJ, AA4OC, WØHI, DD8BA, KN7D, JP1EWY, JG3FEA, WØBM, IZ8GUQ, 9HS1P, NZ3D
Oceania: JR3UIC, WF2S, KN7D, AB1OC, JP1EWY, JG3FEA, 9H1HP
North America: N6ETO, KI7DG, K1BZ, KK4FYV, R3LC, N3FHW, JR3UIC, KC8UR, WØHI, AC7JM, WB9G, KN7D, JG3FEA, WØBM, 9H1SP
South America: 9H1SP

Award of Excellence:
Award of Excellence with 160 Bar: AB1OC, 9H1SP
160M Bar: K6SIK
30M Bar: K6SIK, 9H1SP
17M Bar: 9H1SP
12M Bar: K6SIK, 9H1SP
6M Bar: 9H1SP
Digital Bar: AB1OC

Award of Excellence Holders: N4MM, W4CRW, K5UR, K2VV, VE3XN, DL1MDD, DJ7CX, DL3RK, WB4SIJ, DL7AA, ON4QX, 9A2AA, OK3EA, OK1MP, N4NO, ZL3GO, W4BQY, IØJX, WA1JMP, KØJN, W4VQ, KF2O, WB8CNL, W1JR, F9RM, W5UR, CT1FL, WA4QMQ, W8ILC, VE7DP, K9BG, W1CU, G4BUE, N3ED, LU3YL/W4, NN4Q, KA3A, VE7WJ, VE7IG, N2AC, W9NUF, N4NX, SMØDJZ, DK5AD, WD9IIC, W3ARK, LA7JO, VK4SS, I8YRK, SMØAJU, N5TV, W6OUL, WB8ZRL, WA8YTM, SM6DHU, N4KE, I2UIY, I4EAT, VK9NS, DEØDXM, DK4SY, UR2QD, AB9O, FM5WD, I2DMK, SM6CST, VE1NG, I1JQJ, PY2DBU, HI8LC, KA5W, K3UA, HA8UB, HA8XX, K7LJ, SM3EVR, K2SHZ, UP1BZZ, EA7OH, K2POA, N6JV, W2HG, ONL-4003, W5AWT, N3XX, HB9CSA, F6BVB, YU7SF, DF1SD, K7CU, I1POR, K9LJN, YBØTK, K9QFR, 9A2NA, W4UW, NXØI, WB4RUA, I6DQE, I1EEW, I8RFD, I3CRW, VE3MS, NE4F, KC8PG, F1HWB, ZP5JCY, KA5RNH, IV3PVD, CT1YH, ZS6EZ, KC7EM, YU1AB, IK2ILH, DEØDAQ, I1WXY, LU1DOW, N1IR, IK4GME, VE9RJ, NN1N, HB9AUT, KC6X, N6IBF, W5ODD, IØRIZ, I2MQP, F6HMJ, HB9DDZ, WØULU, K9XR, JAØSU, I5ZJK, I2EOW, IK2MRZ, KS4S, KA1CLV, WZ1R, CT4UW, KØIFL, WT3W, IN3NJB, S50A, IK1GPG, AA6WJ, W3AP, OE1EMN, W9IL, I7PXV, S53EO, DF7GK, S57J, EA5BM, DL1EY, DJ1YH, KUØA, VE2UW, 9A9R, UAØFZ, DJ3JSW, OE6CLE, HB9BIN, N1KC, SM5DAC, RW9SG, WA3GNW, S51U, W4MS, I2EAY, RAØFU, CT4NH, EA7TV, W9IAL, LY3BA, K1NU, W1TE, UA3AP, EA5AT, OK1DWC, KX1A, IZ5BAM, K4LQ, KØKG, DL6ATM, VE9FX, DL2CHN, W2OO, AI6Z, RU3DX, WB9IHH, CT1EEN, G4PWA, OK1FED, EU1TT, S53MJ, DL2KQ, RA1AOB, KT2C, UA9CGL, AE5B, KØDEQ, DKØPM, SV1EOS, UAØFAI, N4GG, UA4RZ, 7K3QPL, EW1CQ., UA4LY, RZ3DX, UA3AIO, UA4RC, N8BJQ, UA3BS, UA9FGR, UT3UY, WA5VGI, UT9FJ, UT4EK, K9UQN, UR5FEO, LY2MM, N3RC, OH3MKH, RA3CQ, UT3IZ, S55SL, RU3ZX, YO9HP, RA3DNC, K8ZT, KE5K, JH8BOE, TF8GX, S58MU, UX1AA, AB1J, DM3FZN, AG4W, UA3QNS, RX3AGD, WB5JID, LY3W, LY5W, RW4WZ, VO1CV, VE1YX, DK8MCT, HB9DDO, DL4CW, W9RPM, IZ3ENH, DM2DXA, EY8MM, K4HB, K6ND, TF3Y, K4CN, W1RM, W3LL, 4Z1UF, W3UA, N8VV, HA8QC, LU5OM, US3IZ, RV9CX, K6UM, RWØLT, 4L1MA, UR1MI, IV3ARJ, K6SIK, R3IS, R9MJ, DG7RO.

160 Meter Endorsements: N4MM, W4CRW, K5UR, VE3XN, DL3RK, OK1MP, N4NO, W4BQY, W4VQ, KF2O, W8CNL, W1JR, W5UR, W8ILC, K9BG, W1CU, G4BUE, LU3YL/W4, NN4Q, VE7WJ, VE7IG, W9NUF, N4NX, SMØDJZ, DK5AD, W3ARK, LA7JO, SMØAJU, N5TV, W6OUL, N4KE, I2UIY, I4EAT, VK9NS, DEØDXM, UR2QD, AB9O, FM5WD, SM6CST, I1JQJ, PY2DBU, HI8LC, KA5W, K3UA, K7LJ, SM3EVR, UP1BZZ, K2POF, IT9TQH, N6JV, ONL-4003, W5AWT, N3XX, F6BVB, YU7SF, DF1SD, K7CU, I1POR, K9LJN, YBØTK, K9QFR, W4UW, NXØI, WB4RUA, I1EEW, ZP5JCY, KA5RNH, IV3PVD, CT1YH, ZS6EZ, YU1AB, IK4GME, NN1N, W5ODD, IØRIZ, I2MQP, F6HMJ, HB9DDZ, K9XR, JAØSU, I5ZJK, I2EOW, KS4S, KA1CLV, KØIFL, WT3W, IN3NJB, S50A, IK1GPG, AA6WJ, W3AP, S53EO, S57J, DL1EY, DJ1YH, KUØA, VR2UW, UAØFZ, DJ3JSW, OE6CLD, HB9BIN, N1KC, SM5DAC, S51U, RAØFU, CT4NH, EA7TV, LY3BA, K1NU, W1TE, UA3AP, OK1DWC, KX1A, IZ5BAM, DL6ATM, W2OO, RU3DX, WB9IHH, G4PWA, OK1FED, EU1TT, S53MJ, DL2KQ, RA1AOB, UA9CGL, SM6DHU, KØDEQ, DKØPM, SV1EOS, N4GG, UA4RZ, 7K3QPL, EW1CQ, UA4LY, RZ3DX, UA3AIO, UA4RC, N8BJQ, UA3BS, UA9FGR, UT3UY, WA5VGI, UT9FJ, UR5FEO, N3RC, UT3IZ RU3ZX, YO9HP, RA3DNC, K8ZT, KE5K, JH8BOE, S58MU, UX1AA, DM3FZN, AG4W, UA3QNS, RX3AGD, LY3W, LY5W, VO1CV, HB9DDO, DL4CW, W9RPM, IZ3ENH, DM2DXA, EY8MM, K4HB, K6ND, W1RM, W3LL, 4Z1UF, W3UA, RV9CX, K6UM, UR1MI, IV3ARJ, R3IS, R9MJ, DG7RO, K0DEQ.

Complete rules and application forms may be obtained by sending a business-size, self-addressed, stamped envelope (foreign stations send extra postage for airmail) to "CQ WPX Awards," P.O. Box 355, New Carlisle, OH 45344 USA. Note: WPX will now accept prefixes/calls which have been confirmed by eQSL.cc. and the ARRL Logbook of The World (LoTW).

*Please Note: The price of the 160, 30, 17, 12, 6, and Digital bars for the Award of Excellence are $6.50 each.

Indeed, some rethinking has been occurring. The organizers of another DXpedition to such a southern spot have concluded that the expense for transportation is absolutely excessive. These folks are now considering a stripped-down effort with limited exposure, a limited number of operators and number of QSOs, and—most importantly—a greatly reduced budget. The thinking is that an expedition that can give everyone one to three contacts is better than no expedition, and certainly better than QSOs at a cost of near $15 or even $20 per QSO. All of this sug-

5 Band WAZ

As of December1, 2013 1850 stations have attained at least the 150-zone level, and 911 stations have attained the 200-zone level.

New recipeints of 5 Band WAZ with all 200 zones confirmed:

DL5AN

The top contenders for 5 Band WAZ (zones needed on 80 or other if indicated):

N4NX, 199 (26)	RWØLT, 199 (2 on 40)
KØQC, 199 (26)	JA1CMD, 199 (2)
W4DC, 199 (24)	I5REA, 199 (31)
N4WW, 199 (26)	RZ3EC, 199 (1 on 40)
W4LI, 199 (26)	W1FJ, 199 (24)
K7UR, 199 (34)	W6OUL, 198 (37,40)
IK8BQE, 199 (31)	EA5RM, 198 (1,19)
JA2IVK, 199 (34 on 40)	N8LJ, 198 (17,24)
IK1AOD, 199 (1)	EA5BCX, 198 (27,39)
VO1FB, 199 (19)	G3KDB, 198 (1,12)
KZ4V, 199 (26)	JA1DM, 198 (2,40)
W6DN, 199 (17)	9A5I, 198 (1,16)
W3NO, 199 (26)	G3KMQ, 198 (1, 27)
RU3FM, 199 (1)	N2QT, 198 (23,24)
N3UN, 199 (18)	OK1DWC, 198 (6, 31)
W1FZ, 199 (26)	W4UM, 198 (18,23)
SM7BI, 199 (31)	US7MM, 198 (2,6)
EA7GF, 199 (1)	K2TK, 198 (23, 24)
JA5IU, 199 (2)	K3JGJ, 198 (24,26)
RU3DX, 199 (6)	F5NBU, 198 (19,31)
N4XR, 199 (27)	W9XY, 198 (22,26)
HA5AGS, 199 (1)	KZ2I, 198 (24,26)
JH7CFX, 199 (2)	W9RN, 198 (26,19 on 40)
RA6AX, 199 (6 on 10)	W5CWQ, 198 (17,18)
RX4HZ, 199 (13)	UA4LY, 198 (6 and 2 on 10)
S58Q, 199 (31)	JA7XBG, 198 (2 on 80 & 10)
K8PT, 199 (26)	JA3GN, 198 (2 on 80 & 40)
N8AA, 199 (23)	N4GG, 198 (18,24)
IZ1ANU, 199 (1)	K4JLD, 198 (18,24)
IN3ZNR, 199 (1)	NS6C, 198 (17,22)
JK1BSM, 199 (2)	

The following have qualified for the basic 5 Band WAZ Award:

N1WQ (172 zones)	9A2NO (172 zones)
ON6KE (150 zones)	KB4CL (163 zones)
DL7KL (170 zones)	

5 Band WAZ updates:

ON4CAS(194 zones)	LU5OM (200 zones)
F4GTB (173 zones)	

***Please note: Cost of the 5 Band WAZ Plaque is $100 shipped within the U.S.; $120 all foreign (sent airmail).**

Rules and applications for the WAZ program may be obtained by sending a large SAE with two units of postage or an address label and $1.00 to: WAZ Award Manager, Floyd Gerald, N5FG, P.O. Box 449, Wiggins, MS 39577-0449. The processing fee for the 5BWAZ award is $10.00 for subscribers (please include your most recent *CQ* mailing label or a copy) and $15.00 for nonsubscribers. An endorsement fee of $2.00 for subscribers and $5.00 for nonsubscribers is charged for each additional 10 zones confirmed. Please make all checks payable to Floyd Gerald. Applicants sending QSL cards to a *CQ* checkpoint or the Award Manager must include return postage. N5FG may also be reached via e-mail: <n5fg@cq-amateur-radio.com>.

gests that we might need to reconfigure the thinking that every DXpedition must cater to the needs of those chasing the DXCC Challenge, and even more so, the competition engendered by the Club Log phenomenon. Is such extravagance good for DXing? Increasingly, we are just saying "No!"

By the time this column appears in print, we will know how the planned January DXpedition to Amsterdam Island ended. In early December, the Amsterdam Island webpage is stating that if they make 100,000 QSOs, each QSO will cost about $4.00. Although not the $10 to $20 per QSO that is envisioned for larger expeditions, it is still excessive. Budgets like that beg the question of whether we are thinking of all possible scenarios. I have said that such efforts will continue as long as they can be funded. Will such funding continue into the future? Or will DXers say "enough!"? Are there alternatives? The DX University will begin studying the issue in the coming months. If you have suggestions and alternative scenarios, please visit the DX University website and register your thoughts in the "Contact Us" section.

July – Remote-Control DXing

For July, I wrote about remote-control DXing. This topic has been simmering for some time. I must have been prophetic, because the ARRL Board of Directors resolved in July to ask the League's DX Advisory Committee to "…study and, if warranted, recommend changes to the DXCC Rules for the ARRL Board of Directors to discuss." There are probably two primary issues: One is that DXCC rules now rely primarily on the whims of the United Nations, which is currently paralyzed on some issues by the prospect of a veto. As a result, it appears that unless the DXCC rules are augmented, it will be a long time before Kosovo—*and other potential entities*—count for DXCC. Since this was published, we have come to further understand that *no ITU prefix will be forthcoming any time soon*. Hopefully, the DXAC and the BoD will grasp what this means and act accordingly. I will reserve judgment.

The other major issue is remote control of amateur radio stations. Through advances in Internet-based devices, it is simple and inexpensive to control your station from anywhere in the world. Many hams don't understand just how extensive the possibilities are: remote control of whole stations, control of receivers on other continents, control of mega-stations owned by others, virtually anything you can think of, and more. Remote control is not in the future. It is *now*. It's legal in most cases, and rules that attempt to regulate or restrict it are totally unenforceable.

August – DXpedition Fairness: "Who to Work"

The August column dealt with fairness by DXpeditions in working all areas of the world. If a DXpedition operates from the Indian Ocean, DXers in the western USA might have to scramble to make even a single contact. Low-band contacts to that area are even more difficult. The difficulty in making QSOs between relatively remote locations can be alleviated by DXpedition operators listening at the proper times for certain locations. This isn't always easy to do, but operations in recent months have done a reasonably good job in this respect.

September – Back in the Day, Some DXing History

The September column was something else! I just look at the picture of KN6ALH (me) in 1953 and shake my head. Certainly lots of DX has come and gone since then. Such a way to measure your life! I guess to stay young we need to keep making DX.

October –The DXPP

Closely tied to the matter of DXpedition funding is that of the DXPP, the *DXpedition Paying Passenger.* The term DXPP has been viewed by some as derogatory, but I believe it is important to first understand what it means. The DXPP is an amateur with (1) time, (2) resources, and especially (3) a good attitude and a willingness to learn the ropes. Of course, the DXPP might not even be interested in operating, but perhaps is simply interested in traveling to a remote location, or possibly just observing the whole thing. Unfortunately, if inadequately prepared or supervised, the DXPP can be a serious liability for a DXpedition. These operators are not capable of running pile-ups efficiently, can give the expedition a bad name, and among other negatives, make future fund raising difficult.

Since October's column I have learned another term. It's the *Inverse*

The WAZ Program

12 Meters CW

78 .....LU5OM 79 .....W6ENZ

17 Meters CW

102 .....LU5OM

30 Meters CW

122 .....LU5OM

80 Meters CW

97 .....DL4MO

160 Meters

435.....UT5IM (40 zones) 436.....R3VA (31 zones)

160 Meter Updates

N5FG (33 zones)

All Band WAZ
Mixed

9068 .....PA1NL 9072 .....AB4SF
9069 .....IZ8FTW 9073 .....N4QWF
9070 .....JL1BDI 9074 .....OZ4VW
9071 .....RN4CA

SSB

5279.....VK2SF

CW

737.....US3LR 739 .....K8BXD
738 .....K5ZE 740.....DD5KG

Satellite

028 .....AHØU (32 zones)

Rules and applications for the WAZ program may be obtained by sending a large SAE with two units of postage or an address label and $1.00 to: WAZ Award Manager, Floyd Gerald, N5FG, P.O. Box 449, Wiggins, MS 39577-0449. The processing fee for all *CQ* awards is $6.00 for subscribers (please include your most recent *CQ* mailing label or a copy) and $12.00 for nonsubscribers. Please make all checks payable to Floyd Gerald. Applicants sending QSL cards to a *CQ* checkpoint or the Award Manager must include return postage. N5FG may also be reached via e-mail: <n5fg@cq-amateur-radio.com>.

CQ DX Awards Program

CW

1150 .....WØOVM

SSB

2617 .....WØOVM 2618.....UX7QC

Endorsements

WØOVM.....(SSB 196)

The basic award fee for subscribers to *CQ* is $6. For non-subscribers, it is $12. In order to qualify for the reduced subscriber rate, please enclose your latest *CQ* mailing label with your application. Endorsement stickers are $1.00 each plus SASE. Updates not involving the issuance of a sticker are free. All updates and correspondence must include an SASE. Rules and application forms for the CQ DX Awards may be found on the <www.cq-amateur-radio.com> website, or may be obtained by sending a business-size, self-addressed, stamped envelope to CQ DX Awards Manager, Please make checks payable to the Award Manager, Keith Gilbertson. Mail all updates to Keith Gilbertson, KØKG, 21688 Sandy Beach Lane, Rochert, MN 56578-9604 USA. We recognize 341 active countries. Please make all checks payable to the award manager. Photocopies of documentation issued by recognized national Amateur Radio associations that sponsor international awards may be acceptable for CQ DX award credit in lieu of having QSL cards checked. Documentation must list (itemize) countries that have been credited to an applicant. Screen printouts from eQSL.cc that list countries confirmed through their system are also acceptable. Screen printouts listing countries credited to an applicant through an electronic logging system offered by a national Amateur Radio organization also may be acceptable. Contact the CQ DX Award Manager for specific details.

DXPP. This individual is such an expert at certain aspects of DXpeditioning, working exclusively on 160 meters, for example, that he is encouraged to come along on a DXpedition on a reduced or no-cost basis. Essentially, the remainder of the participants or even the foundations support these people because of their expertise. The Inverse DXPP.

November – The DXAC's New Assignment

Following up on my July premonition, I noted in November that the ARRL BoD asked the DXAC to study several issues. In fact, the board asked the DXAC to review *all* DXCC rules and report back. By the time you read this, the committee will have written an interim report, which most likely will have been reviewed by the ARRL Board at its January meeting. By the time this appears, we may well be hearing of progress by the DXAC.

December – DXing with a Modest Station

In the December column, I discussed how one might work DX with a modest station. When I solicited input, I was universally asked, "How do you define a modest station?" Generally, I consider a modest station to be one running 100 watts with an antenna that will fit on a small city property. Since hearing from some DXers who are running such stations, I find that they do very well, beating me out in many pile-ups. (Maybe my station is more modest than I thought.) What that says is that waiting for better or the best propagation will win out every time. Since this was written, we have had more excellent conditions, more sunspots. Perhaps this is the answer for those with modest stations.

There you have it, my thoughts from April through December 2013 with some updates. What will 2014 bring?

Missionary DXpeditioning from South Sudan

The mission of the November DXpedition to South Sudan, Z81X, was two-fold. First, it was to present a workshop on amateur radio to the Ministry of Telecommunications in Juba. While I don't have the space here to discuss this fully, I can say that a whole day was spent with the Director General of the Ministry and 20 other members of that administration. A presentation was made by Hans Blondeel Timmerman, PB2T, president of the IARU, Region 1; Martti Laine, OH2BH; and your columnist, with information on (1) Training regulators in the administration of the amateur service; (2) Assisting the administration in the creation of amateur radio regulations and related frequency table; (3) Discussing amateur radio's benefits for the society; and (4) Evaluating the prospects of establishing the amateur radio service for South Sudanese nationals. Several pictures taken at this workshop accompany this column.

The DXpedition Operating from South Sudan

While it's still fresh in my mind, I also want to make some observations about the operating I observed while operating Z81X from Juba, South Sudan at the end of November 2013.

CQ DX Honor Roll

The CQ DX Honor Roll recognizes those DXers who have submitted proof of confirmation with 275 or more ACTIVE countries. With few exceptions, the ARRL DXCC Countries List is used as the country standard. The CQ DX Award currently recognizes 341 countries. Honor Roll listing is automatic when an application is received and approved for 275 or more active countries. Deleted countries do not count and all totals are adjusted as deletions occur. To remain on the CQ DX Honor Roll, annual updates are required. All updates must be accompanied by an SASE if confirmation of total is required. The fee for endorsement stickers is $1.00 each plus SASE. (Stickers for the 340 level are available.) Please make checks payable to the Award Manager, Keith Gilbertson. Mail all updates to Keith Gilbertson, KØKG, 21688 Sandy Beach Lane, Rochert, MN 56578-9604 USA.

CW

OH2BN.....341	K4IQJ.....340	K4CN.....340	K9IW.....339	W6OUL.....336	K6LEB.....331	KA3S.....325	WA4DOU.....313	N3RC.....295
NØFW.....340	K4MQG.....340	K9BWQ.....340	W7OM.....339	PY2YP.....334	N7WO.....331	EA5BY.....325	YO9HP.....313	HA1ZH.....292
K2TQC.....340	W8XD.....340	WØJLC.....340	YU1AB.....339	K9OW.....334	W1DF.....331	OZ5UR.....325	N2LM.....313	WA2VQV.....285
WB4UBD.....340	N7RO.....340	W7CNL.....340	N4CH.....339	K1FK.....334	WG5G/	K8ME.....323	HB9DAX/	K7CU.....282
K3UA.....340	N5FG.....340	N4MM.....340	K7LAY.....339	K2OWE.....333	QRPp.....330	WD9DZV.....323	QRP/p.....311	PP7LL.....282
N4JF.....340	K4CN.....340	K9IW.....340	K3JGJ.....339	W4MPY.....333	JA7XBG.....329	ON4CAS.....322	AB4IQ.....310	N2VW.....280
K2FL.....340	OK1MP.....340	K5RT.....340	K8LJG.....338	K5UO.....333	K6YK.....329	KØKG.....320	KT2C.....306	K4EQ.....280
WK3N.....340	N4AH.....340	W5BOS.....340	KA7T.....338	N6AW.....333	IKØADY.....328	W6YQ.....319	K7ZM.....304	4Z5SG.....276
W4OEL.....340	F3TH.....340	N4NX.....339	W7IIT.....337	HB9DDZ.....333	W9IL.....327	CT1YH.....317	HA5LQ.....301	YO6HSU.....275
EA2IA.....340	DL3DXX.....340	K4JLD.....339	K8SIX.....337	K7VV.....333	KE3A.....326	EA3ALV.....316	K8IHQ.....301	
N7FU.....340	K9MM.....340	N5ZM.....339	G3KMQ.....337	F6HMJ.....332	K6CU.....326	RA1AOB.....314	K4IE.....295	

SSB

XE1AE.....341	K4IQJ.....341	W7BJN.....341	W7OM.....340	W2CC.....339	N6AW.....334	WØROB.....329	AD7J.....317	K4ZZR.....305
NØFW.....341	N5FG.....341	AB4IQ.....341	K9HQM.....340	W9IL.....338	IK8CNT.....334	K7HG.....328	WD9DZV.....315	K7ZM.....304
K6YRA.....341	K4CN.....341	VE1YX.....341	KØKG.....340	K3LC.....337	CT3BM.....334	KD5ZD.....328	HB9DQD.....314	4Z5FL/M.....303
IK1GPG.....341	OZ3SK.....341	K8SIX.....341	YU1AB.....340	VK2HV.....336	K8LJG.....334	W1DF.....327	N8SHZ.....313	K7SAM.....302
K2TQC.....341	OK1MP.....341	W8ILC.....341	VE3MR.....340	W4WX.....336	HB9DDZ.....334	KE4SCY.....327	IV3GOW.....313	KA8YYZ.....302
K4MZU.....341	N4CH.....341	K5TVC.....341	VE3MRS.....340	IØZV.....337	KE3A.....333	K6GFJ.....327	W6NW.....311	N3RC.....301
DJ9ZB.....341	DL3DXX.....341	I8KCI.....341	K7LAY.....340	OE2EGL.....336	N2VW.....333	W9GD.....326	KU4BP.....311	4X6DK.....298
WB4UBD.....341	OZ5EV.....341	W5BOS.....341	K7VV.....340	EA3BMT.....336	JA7XBG.....333	VE7SMP.....326	AE9DX.....311	K2HJB.....295
N4JF.....341	DU9RG.....341	K5RT.....341	W4UNP.....340	IKØAZG.....336	N5YY.....333	N2LM.....326	I3ZSX.....310	VE6MRT.....296
WK3N.....341	N4MM.....341	K5OVC.....341	K9IW.....340	W8AXI.....335	K5UO.....332	KF4NEF.....326	KA1LMR.....309	W9ACE.....291
K4JLD.....341	KE5K.....341	VE2PJ.....341	K8LJG.....339	VK4LC.....335	K8ME.....332	VE7EDZ.....325	RA1AOB.....309	W6MAC.....289
N7BK.....341	IN3DEI.....341	W4ABW.....341	4Z4DX.....339	WS9V.....335	W6OUL.....332	F6BFI.....324	G3KMQ.....309	N3KV.....289
EA2IA.....341	EA4DO.....341	K3UA.....340	W7FP.....339	AA4S.....335	WØYDB.....331	ON4CAS.....324	XE1MEX.....309	K7CU.....287
K4MQG.....341	YU3AA.....341	N4NX.....340	N7WR.....339	PY2YP.....335	WA4WTG.....331	W4MPY.....323	IØYKN.....307	WD8EOL.....281
K9MM.....341	K9BWQ.....341	VE3XN.....340	W2FKF.....339	K9OW.....335	ZL1BOQ.....331	TI8II.....321	XE1MW.....306	IZ1JLG.....282
K3JGJ.....341	KZ2P.....341	W6BCQ.....340	W3AZD.....339	EA5BY.....335	SV3AQR.....331	YO9HP.....321	AA1VX.....305	IWØHOU.....277
N5ZM.....341	W9SS.....341	K2FL.....340	K1UO.....339	XE1J.....335	CT1AHU.....329	KW3W.....321	W5GT.....305	WA5UA.....276
N7RO.....341	VE2GHZ.....341	W6DPD.....340	F6HMJ.....339	OE3WWB.....334	N1ALR.....329	XE1RBV.....318	K4IE.....305	SQ7B.....275

RTTY

WB4UBD.....339	N5FG.....337	OK1MP.....335	WK3N.....334	K8SIX.....325
NI4H.....339	N5ZM.....336	K4CN.....334	K3UA.....332	AB4IQ.....295

The South Sudan workshop group, including Hans, PB2T; Martti, OH2BH; and Wayne, N7NG. See text for details.

Much has been said in recent years about the deteriorating operating ability and skills of the world's DXers. Overall, my experience doesn't support much of what I consider so much hand-wringing.

To start, I have to admit that my own operating revealed some roughness around the edges, particularly at the beginning. Besides having spent considerable time recently DXing rather than DXpeditioning and running pile-ups, dealing with unfamiliar equipment took its toll. Imagine sitting down at a station and realizing that the keyboard is in German (or Spanish) and that not all of the keys are in the same position as I was accustomed to. I did have my own computer with the same software, but changing them between shifts seemed initially more of a pain that it was worth.

The actual operating was lots of fun. On only a few occasions did I find the pile-up out of control, with stations calling when they should have been listening. Most of this was understandable, and when it was necessary to regain control, repeating a partial callsign several times proved sufficient.

There were only a few instances in

The workshop attendees were treated to a catered lunch. (OH2BH photo)

which I had to deal with stations calling out of turn to the point that I couldn't copy the desired station. There were also a few situations in which my frequency was being QRMed by stations calling me on the wrong (my) frequency. I did note a few of these cases where my rate suffered—because people couldn't hear me due to the QRM on my frequency—and I had to change my frequency by a few hundred Hertz. Generally, however, my rate was up to standards, indicating that the pile-up management was satisfactory. (Note that it doesn't really matter what *you* think of the QRM on my frequency. It's what *I* hear that is important.) Before we terminated the operation, I asked some of the operators (OHØXX, OH6KN, DL3DXX) what they thought of the operating in the pile-ups. Generally, their impressions were positive. They had no significant complaints. My own experiences echo their observations.

Next month I'll talk a little more about the South Sudan trip and also a little about an out-of-the-box possibility for organizing QSOs on an extreme DXpedition.

73, de Wayne, N7NG

Hans, PB2T, addresses the workshop group. (OH2BH photo)

The Director General of the Ministry of Telecommunications and Postal Services at the front of the class! (OH2BH photo)

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BY GEORGE TRANOS,* N2GA

Can Contesting be Successful as a Spectator Sport?

Anyone is allowed to compete in amateur radio contests, regardless of their previous experience, skill level, or equipment. Get on the radio, put out a signal, make contacts, and you're part of the event! Even the least-experienced novice operator can make contact with the best of the best. As a matter of fact, the best operators *want* to interact with everyone—novices through experts.

So why doesn't everyone contest? Part of what intimidates some people about contesting may be the pages of rules and the seeming complexity of top-notch stations. That can also keep a spectator from appreciating the operating skills, decision-making, and effortlessness of the best contesters. Wouldn't it be nice to see and hear the best operators in the world right from their operating location? Can this be done in a way that allows contesting to become a successful *spectator sport* as well as a participation sport?

You can compare an amateur radio contest (also called *radiosport*) to a major sporting event. Watching sports is big business. Fans around the world spend many hours attending, listening to, and watching sporting events on television and the Internet. These events attract major advertisers who pay large fortunes to have their commercials play during them. Major sports are covered by media worldwide and produce some of the largest viewing audiences. The excitement created during and leading up to spectator sporting events is looked forward to by a major percentage of the world's population. Amateur radio competition should be no different, as it offers its own brand of excitement.

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Similarities

An amateur radio contest has a finite length of time, the contest period. This is equivalent to a "game"in any other sport. There is a distinct winner, along with second-, third-, and fourth-place achievements. The rules are available for anyone to follow. There are periods of great activity, such as a huge pile-up with many operators calling in at once, leading to high rates of contact. There are technological challenges similar to other sports, such as bobsledding or sailing, in which sophisticated software modeling contributes to maximizing the performance of the necessary equipment.

Operators from all continents and hundreds of countries participate in worldwide amateur radio contests. Contest spectators may want to cheer for operators from their country (or state or radio club, etc.), just like fans want to root for their home team. Hams may personally know and like some of the participants and this may form a bond for the spectator as well.

In most contests, operators from many different categories operate all at the same time but are only

Calendar of Events

All year	**CQ DX Marathon http://bit.ly/vEKMWD**
Feb. 1	AGCW Straight Key Party http://bit.ly/1dWd6V9
Feb. 1	Minnesota QSO Party http://bit.ly/1gho8b1
Feb. 1–2	Vermont QSO Party http://www.ranv.org/vtqso.html
Feb. 1–2	Delaware QSO Party http://www.fsarc.org/qsoparty/qsohome.htm
Feb. 1–2	Mexico RTTY Int'l Contest http://www.fmre.org.mx/
Feb. 1–2	10-10 Int'l Winter Contest http://bit.ly/yTsaDk
Feb. 1–2	Black Sea Cup Int'l http://www.bscc.in/index/rules_black_sea_cup_international_eng/0-21
Feb. 2	North American CW Sprint http://www.ncjweb.com/sprintrules.php
Feb. 8	Asia-Pacific Spring Sprint (CW) http://jsfc.org/apsprint/aprule.txt
Feb. 8	FISTS Winter Sprint http://www.fists.org/operating.html
Feb. 8–9	**CQ WW RTTY WPX Contest http://bit.ly/uYC0gp**
Feb. 8–9	Dutch PACC Contest http://pacc.veron.nl/
Feb. 8–9	New Hampshire QSO Party http://www.w1wqm.org/nhqso/
Feb. 8–9	SARL Field Day Contest http://bit.ly/H0IqQf
Feb.10–14	ARRL School Club Roundup http://www.arrl.org/school-club-roundup
Feb. 15–16	ARRL CW DX Contest http://www.arrl.org/arrl-dx
Feb. 21–23	**CQ WW 160M SSB Contest http://www.cq160.com/rules.htm**
Feb. 22–23	REF SSB Contest http://concours.ref-union.org/contest/?page_id=2
Feb. 22–23	UBA CW DX Contest http://bit.ly/W0gZiE
Feb. 22–23	North American RTTY QSO Party http://bit.ly/GLPfXz
Feb. 23	High Speed Club CW Contest http://www.highspeedclub.org/
Mar. 1–2	ARRL SSB DX Contest http://www.arrl.org/arrl-dx
Mar. 4	AGCW YL-CW QSO Party http://bit.ly/T0fKh9
Mar. 8	AGCW QRP Contest http://bit.ly/UKwb0L
Mar. 30–31	**CQ WW WPX SSB Contest http://bit.ly/hKqJjG**

competing against rivals in their own category. This is similar to some types of auto racing, in which different car classes drive on the same racetrack but only vie for positions within their own class.

Differences

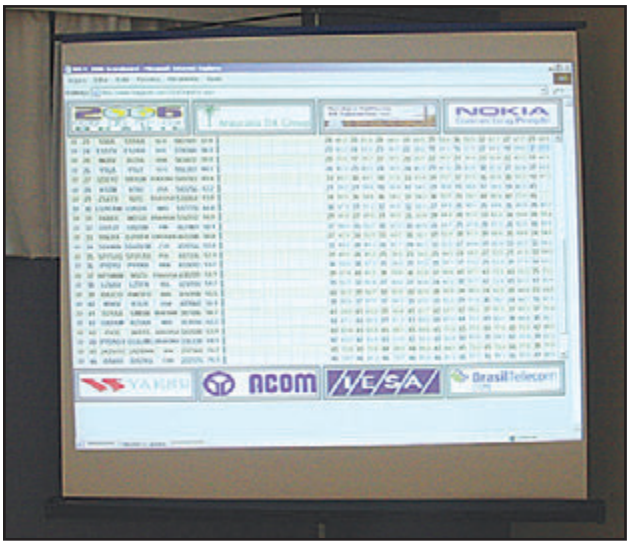
Contesters are geographically distributed and may be anywhere in the world. It may be difficult for a spectator to "watch" the competition from any one spot. When the contest is over, the raw scores can be compared to determine the winner, but the real winner is not known until log checking is performed. That might be "good enough" for the spectator, but the final results may show a different order of finish from the raw score, and that process takes several months.

Ingredients Needed for Success

When you're watching your favorite sports team, don't you want to know the background of the players? In a contest, wouldn't it be great if a spectator could know who the radio operators are and some background information about them personally as well as their station's information? The observer should be able to view the competition as it takes place in real-time, as if they were viewing the contest at each participant's location. There should be a way to compare the score and activity of each operator to other similar stations.

A necessary component for someone to observe and analyze the competition as a spectator would be a scoreboard. The current contest and time remaining should be shown. Without the ability to see the whole event at once, a viewer should be able to see a leaderboard with the competitors listed in order from highest to lowest. To make things more interesting, this scoreboard should show information right from the operators' logging programs. The callsign used, category entered, and current score are the basics. Other data that might make viewing more interesting include the current QSO rate for the last 10 and last 100 contacts; hourly breakdown of score, number of contacts, and multipliers; current active band; station and frequency of last contact.

There are existing websites such as getscores.org, Club Log, and others that provide a basis for just such a scoreboard. Computer program interfaces already exist between most contest logging programs and these websites. This is the key to real-time score reporting. As each contact is logged, it should update the scoreboard so spectators can see the latest score.



The scoreboard at the World Radiosport Team Championship 2006 in Brazil. Note listings for the placement and operator callsigns of each team, total score, and other information. (Photos courtesy of Diane Ortiz, K2DO)

Virtual Visiting

A nice option would be for a spectator to click on one of the callsigns to actually view and hear the operation currently in progress from that station. Video and audio could then be pulled up for the selected operator. For CW and RTTY contesting, a visual display of the on-air audio could show the contest operation as it takes place. Both sent and received information should be shown so the spectator could watch the contest in its entirety. Other logging windows as displayed on the operator's screen should also be available to allow the fan to view the operator display and be "in the cockpit" with the operator. The spectator should then have the option of switching to a competitor's station. All the same options would be available. Other operators should have the option of determining if they're in the log. This would allow others in the contest to participate and "spectate" at the same time!

Some other things that might create excitement would be the ability to compare callsigns. The viewer could select, say, up to three stations, and current data would be shown on the screen side-by-side. This real-time information would change every time a contact is logged. Comparative statistics would show which station is growing its QSO rate and which is slowing.

Other interesting items for viewers would be the station and antenna setup. Where is the station located? Is it on a mountaintop with views in all directions or is the operation on a small city lot? Station setup is a big part of contesting

The scoreboard at WRTC 2006 was visible at the headquarters hotel as well as over the Internet. People gathered to view the contest as it was taking place.

today, but operator ability distinguishes the top scorers from the also-rans. Some great operators do a fantastic job from a small station but can't compete with the leaders because the hardware is just not there. Knowing this helps the spectator put the station performance in perspective.

Statistics are other areas where fans glean information and can debate about sports. Contesters have histories that are available for comparison. How did this station do last year or the year before in the same contest? How does that compare with the second- or third-place operator? The more information there is available, the more interesting the contest can become to a fan.

Summary

Contesting is an exciting sport but it's hard to demonstrate that excitement to non-contesters. If we can find a way to highlight our sport and actually show it in operation, we could perhaps get more hams interested in contesting. We might also be able to entice the general public into becoming ham radio operators and contesters. Kids today are really interested in gaming and online competition, and ham radio contests are very similar. If we could find a way to show them how exciting ham radio contesting can be, then they might want to get involved.

Could we get to the point where radiosport is treated like other spectator sports? That might be a stretch, but we currently have the technology to provide a more fan-friendly environment. Anything we can do to highlight the achievements and activity of contesters will go a long way toward promoting our sport and providing new blood for future contests. While contesting is already a growing activity in ham radio, it might expand even faster if non-contesters saw for themselves the fun and camaraderie the sport has to offer.

73, George, N2GA

Wouldn't it be nice to be part of the action by tuning into video and audio right from the contest site? Here's the operation of contest station PT5G by Pat Barkey, N9RV, and Tim Duffy, K3LR, during the World Radiosport Team Championship 2006. It took place at the station of PP5JR in Sambaqui, Brazil.

WRTC 2006 had stations spread out throughout the southern portion of Brazil. With Florianopolis as its headquarters, stations were distributed up and down the coast."

Looking Ahead in CQ *and* CQ plus

Here are some of the articles we're working on for upcoming issues of *CQ*:

- **Results, 2013 CQ World-Wide RTTY DX Contest**
- **Results, 2013 CQ World-Wide Foxhunting Weekend**
- The CW Neighborhood
- 3 Scientists … and their Radio Roots

… and in the new CQ-Plus:

- The HI JUNO Project
- "Elmering" in the Internet Age
- Making the Most of AM BCB DXing Tools
- The 100-mW "AM Transmitter Challenge"

Upcoming Special Issues

June: Take it to the Field
October: Emergency Communications
December: Technology

Do you have a hobby radio story to tell? Something for one of our specials? *CQ* now covers listening and personal two-way services in addition to amateur radio. See our writers' guidelines on the CQ website at <http://www.cq-amateur-radio.com/cq_writers_guide/cq_writers_guide.html>.

Equinoctal Propagation

BY TOMAS HOOD,* NW7US

Beginning about the middle of February and continuing through March and early April, typical equinoctial propagation conditions can be expected on the HF frequencies. This usually means a noticeable improvement in conditions between the northern and southern hemispheres. Look for improvements between the United States and South America, Africa, Australasia, Antarctica, and parts of Asia. Equinoctial propagation occurs during the spring and fall months, when the sun is most directly overhead at the equator, producing similar ionospheric characteristics over large areas of the world. It tends to maximize during sunrise and sunset periods and over both short- and long-path openings.

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e-mail: <nw7us@hfradio.org>

A Quick Look at Current Cycle 24 Conditions

(Data rounded to nearest whole number)

Sunspots
Observed Monthly, November 2013: 78
Twelve-month smoothed, May 2013: 60

10.7 cm Flux
Observed Monthly, November 2013: 148
Twelve-month smoothed, May 2013: 118

***Ap* Index**
Observed Monthly, November 2013: 5
Twelve-month smoothed, May 2013: 7

One Year Ago: A Quick Look at Solar Cycle Conditions

(Data rounded to nearest whole number)

Sunspots
Observed Monthly, November 2012: 53
Twelve-month smoothed, May 2012: 65

10.7 cm Flux
Observed Monthly, November 2012: 123
Twelve-month smoothed, May 2012: 126

***Ap* Index**
Observed Monthly, November 2012: 9
Twelve-month smoothed, May 2012: 8

LAST-MINUTE FORECAST

Day-to-Day Conditions Expected for February 2014

	Expected Signal Quality			
Propagation Index.....	(4)	(3)	(2)	(1)
Above Normal: 2-6, 8-12, 14-18, 22-27	A	A	B	C
High Normal: 13, 19-21	A	B	C	C-D
Low Normal: 7	B	C-B	C-D	D-E
Below Normal: 1, 28	C	C-D	D-E	E
Disturbed: N/A	C-D	D	E	E

Where expected signal quality is:
A—Excellent opening, exceptionally strong, steady signals greater than S9.
B—Good opening, moderately strong signals varying between S6 and S9, with little fading or noise.
C—Fair opening, signals between moderately strong and weak, varying between S3 and S6, with some fading and noise.
D—Poor opening, with weak signals varying between S1 and S3, with considerable fading and noise.
E—No opening expected.

HOW TO USE THIS FORECAST

1. Find the *propagation index* associated with the particular path opening from the Propagation Charts appearing in *The New Shortwave Propagation Handbook* by George Jacobs, W3ASK; Theodore J. Cohen, N4XX; and Robert B. Rose, K6GKU.

2. With the *propagation index*, use the above table to find the expected signal quality associated with the path opening for any given day of the month. For example, an opening shown in the Propagation Charts with a *propagation index* of 3 will be poor (D) to fair (C) on Feb. 1st, excellent (A) from the 2nd to the 6th, Fair (C) to good (B) on the 7th, etc..

3. As an alternative, the Last-Minute Forecast may be used as a general guide to space weather and geomagnetic conditions through the month. When conditions are Above Normal, for example, the geomagnetic field should be quiet and space weather should be mild. On the other hand, days marked as Disturbed will be riddled with geomagnetic storms. Propagation of radio signals in the HF spectrum will be affected by these conditions. In general, when conditions are High Normal to Above Normal, signals will be more reliable on a given path, when the path is ionospherically supported.

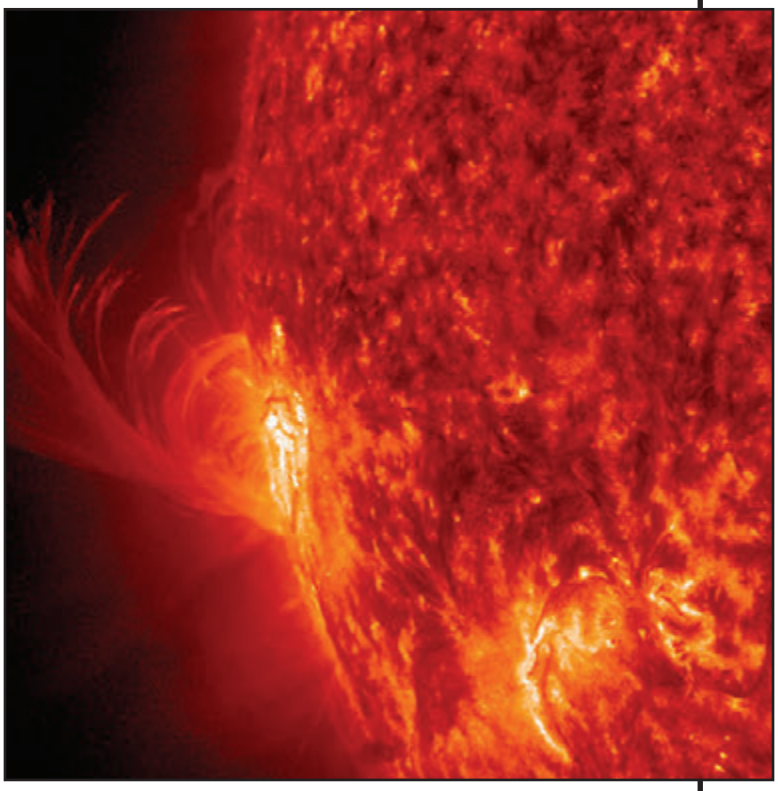


Fig. 1. The Sun burst out with a curling spray of plasma from an active region just coming into view on December 5, 2013. A number of smaller bursts also erupted from the same region, including at least one flare. All of the activity has its source in turbulent magnetic forces associated with the active region, the bright area seen in this image as the bright "patch" out of which the plasma sprays. The video at <http://g.nw7us.us/1dEMN4O>, taken in extreme ultraviolet light, covers about 5.5 hours beginning at 01:00 UTC (notice the x-ray flare, seen as a bright flash). This image was taken at about 02:52 UTC. (Credit: Solar Dynamics Observatory [SDO]/NASA)

During the daylight hours, optimum DX propagation conditions are expected on 15 meters. The band is forecast to open to all areas of the world sometime during this period, often with strong and stable signals with little fading or noise. Openings will be a bit shorter than those of the last few years. Conditions on 10 and 12 meters should run a close second, but with fewer openings expected into Europe and the Far East. Excellent worldwide DX openings to most areas of the world are forecast for 17 and 20 meters during the daylight hours. Conditions are expected to optimize for an hour or two after sunrise and again during the late afternoon. With increasing hours of daylight during February, expect the 10-, 12-, 15-, 17-, and 20-meter bands to remain open for an hour or so longer into the early evening than during the winter months.

Although the solar cycle is meandering along, be sure to check the 6-meter band for possible DX openings, particularly when a number of C-class or higher x-ray flares are occurring, and when the 10.7-cm flux is climbing. Openings are expected to be less numerous than in previous years of higher solar activity, but some openings may still be possible during the hours of daylight. The best bet is for openings toward Central and South America, but other openings may also be possible.

During the early evening hours and to as late as midnight, seven bands should be available for DX openings—15, 17, 20, 30, 40, 80, and 160 meters. Fifteen and 17 meters should hold up for openings toward Central and South America and the Caribbean, the Pacific area, Far East, and parts of Asia. Even better openings to many areas of the world should be possible on 20 meters during this period, with the strongest signals from southerly and westerly directions. Good DX conditions are also forecast for 30, 40, and 80 meters for openings toward the east and the south. Openings in the same direction, but with higher noise levels and weaker signals, should also be possible on 160 meters.

Between midnight and sunrise it should be a toss-up among 20, 30, and 40 meters for DX paths. These bands should open to many areas of the world with conditions favoring openings towards the south and the west. Expect similar conditions on 80 meters, but with weaker signals and higher noise levels. Be sure to check 160 for some unusual DX openings toward the south and the west during this period. Conditions on the bands between 160 and 20

CQ WW CW Contest Conditions

As you may remember, I predicted good to excellent conditions for the 2013 CQ WW CW Contest starting at 0000 UTC Saturday, November 23, and running through 2359 UTC Sunday, November 24. The weekend was, at least by the reports I've received, well within the prediction. The planetary Ap on the first day was a low eight on the first day, falling to a very quiet two by the second day of the contest.

With such quiet geomagnetic conditions, the lower HF bands were great players during the night; weaker signals were easier heard on those bands where openings provided signals weak yet present. During the day, the higher bands played nicely, as well.

The 10.7-cm radio flux was 136 on Saturday and 127 on Sunday, and the observed sunspot count on the first day was 65, while on the second day it was 69. There were ten minor C-class x-ray flares, and two M-class flares on day one. On day two, there were also 10 C-class flares. This activity helped higher-band propagation, but caused a bit of degradation on lower bands. These conditions still provided reasonable propagation on many of the contest bands. These conditions fit pretty well with the forecast.

How did you fare this year compared to last year? I am interested in hearing from you regarding the differences between last year and this year, and how you did overall this time.

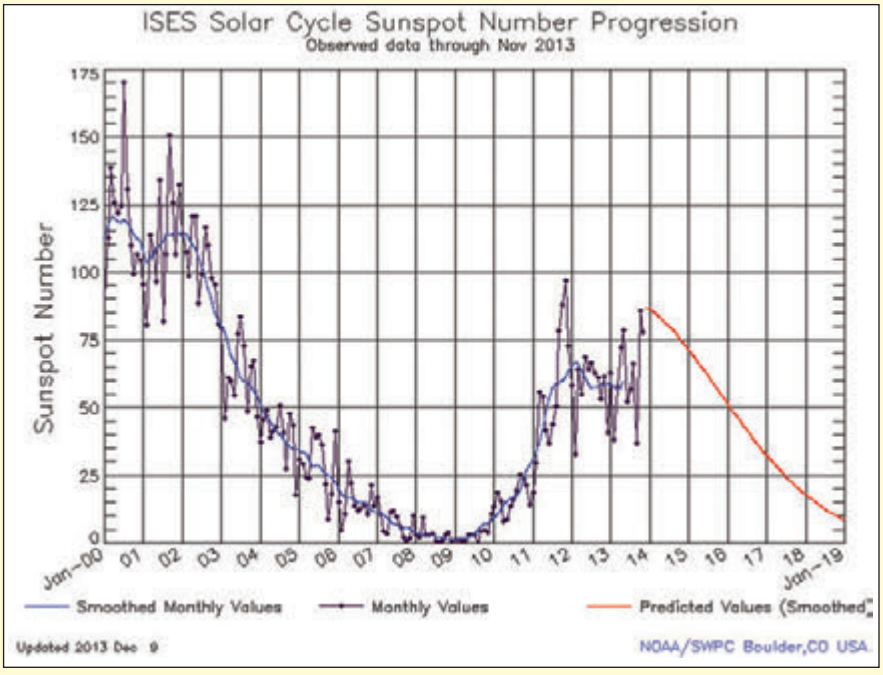


Fig. 2. The plots of current sunspot Cycle 24, as of December, 2013, reveal a significant increase of solar activity during the last months of 2013. While this cycle is much weaker than a number of past solar cycles, this recent uptick in activity has breathed new life in the high-frequncy amateur radio allocations. One might postulate that there is a second peak in Cycle 24. Keep in mind that, magnetically, the Sun is showing signs of reversing polarity (where the north and south solar poles are reversed), a sure sign that we're approaching the down-turn marking the decline phase of the cycle. (Credit: SWPC/NOAA)

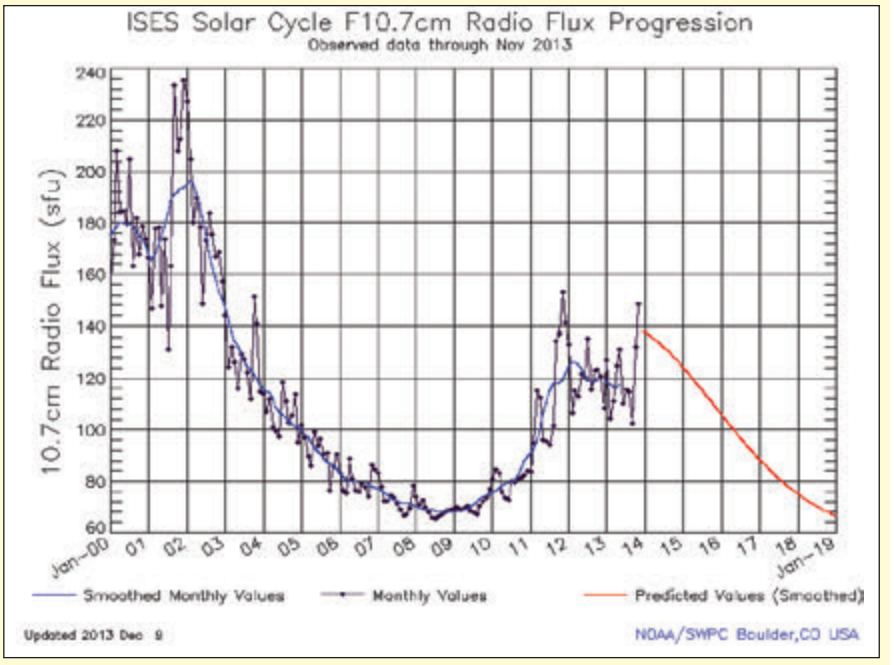
meters are expected to peak at local sunrise.

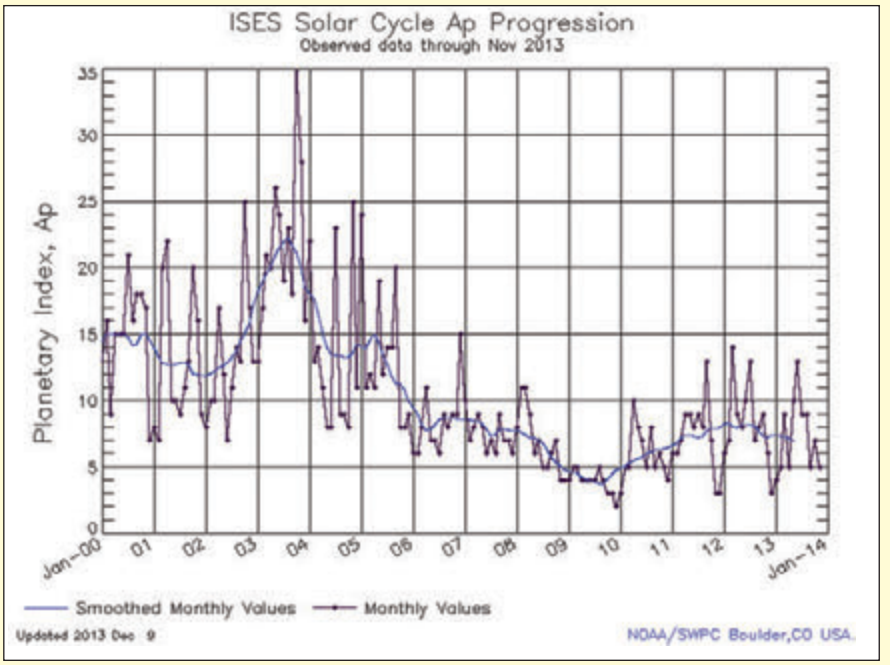
VHF Ionospheric Openings

As mentioned previously, check for 6-meter DX openings during the daylight hours. Some short-skip openings over distances of about 1200 to 2300 miles may also occur. Best times for such openings are during the afternoon hours.

Trans-equatorial (TE) scatter propagation tends to increase during the equinoctial period, and some 6-meter openings may be possible between 7 and 10 PM local time. The best bet for such openings is between the southern tier states and South America for paths approximately at right angles to the equator. An occasional TE opening may also be possible on 2 meters. Unlike F_2-layer or sporadic-*E* openings on 6 meters, TE openings are characterized by very weak signals with considerable flutter fading.

Auroral displays tend to occur somewhat more frequently during the equinoctial period. Unusual short-skip conditions often occur on the VHF bands during these displays. Openings, generally over distances of several hundred and up to about 1300 miles, may take place by means of reflection from





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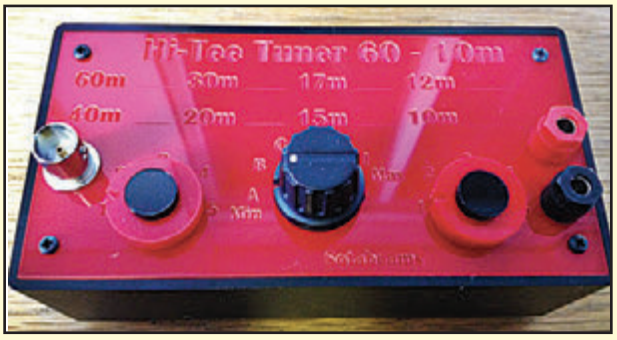
what's new

New Products from SOTAbeams

"Hi Tee" QRP Antenna Tuner

UK-based portable accessory specialists SOTAbeams has introduced the first in its new range of antenna tuners. The Hi Tee Tuner has been designed with performance and weight in mind, according to a company announcement. It weighs just over 5 ounces (150 grams) and covers 60–10 meters.

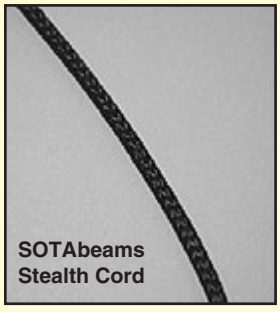
Unlike many QRP tuners, the company says, an air-core coil is used to minimize losses. The tuner has been optimized for portable use, although it is just as much at home in the shack. It includes space on the front to fill in the settings for each band for your favorite antenna. The tuner retails for just over $80US at current exchange rates (plus shipping). For more information, visit <http://www.sotabeams.co.uk/hi-tee-tuner/>.



SOTAbeams "HiTee" QRP Antenna Tuner

SOTAbeams Antenna Cords

Sometimes you want your antenna supports to be as invisible as possible; other times you want high visibility. The folks at SOTAbeams, who specialize in portable antennas, are now offering both options. On the hard-to-see side is *Stealth Cord,* a thin, woven nylon cord consisting of a green and brown weave. Formulated to have low visual impact, it is ideal for holding up stealth antennas. Stealth Cord is supplied on a spool in a generous 50m (160 ft) or 200m (720 ft) lengths, for approximately $12 and $28 US, respectively, plus shipping.


SOTAbeams Stealth Cord

For times when safety is more of an issue than stealth, SOTAbeams also is offering a new "hi viz" cord designed for use on busy hilltops or in emergency situations. It comes in the same lengths at the same prices. Both products are available in SOTAbeams' web-store at <www.sotabeams.co.uk>.

Note: *"What's New" is not a product review and does not constitute a product endorsement by CQ. Information is primarily provided by manufacturers/vendors and has not necessarily been independently verified. The purpose of "What's New" is to inform readers about new products in the marketplace. We encourage you to do additional research on products of interest to you.*

the ionized region produced by an auroral display. Flutter fading and multi-path echoes characterize auroral-type openings. To take maximum advantage of such openings, rotatable antennas should be beamed toward the auroral display, if it is visible.

Large areas of sporadic-*E* ionization also accompany most auroral displays. Reflection of VHF signals from these regions can make possible short-skip openings between distances of 750 and 1300 miles. Signals reflected in this manner are usually strong and stable as compared to those reflected directly from an auroral display.

Auroral activity often occurs during periods of radio storminess on the HF bands. Check the Last-Minute Forecast (also available online at <http:// SunSpotWatch.com>) for those days expected to be Below Normal or Disturbed during February. These are the days on which VHF auroral-type openings are most likely to occur.

Check out *CQ VHF* magazine's Propagation column for an in-depth look at propagation on VHF and above. Be sure to send in your reports to this columnist, as well.

Current Solar Cycle Progress

Canada's Dominion Radio Astrophysical Observatory at Penticton, British Columbia reports a 10.7-cm observed monthly mean solar flux of 148.4 for November 2013, up from 132.3 for October 2013, 46 points higher than September's 102.7! This indicates a clear resurgence of solar activity. This provided a lot of great DX and activity even on the 10-meter band. The 12-month smoothed 10.7-cm flux centered on May 2013 is 118.1, up from April's 116.6. A smoothed 10.7-cm solar flux of about 133 is predicted for February 2014.

The Royal Observatory of Belgium, the world's official keeper of sunspot records, reports a monthly mean sunspot number of 77.6 for November 2013, down a bit from October's 85.6, yet still significantly higher than 36.9 for September 2013. The low for the month was 25 on November 26. The high of 131 occurred on November 17. The mean value for November results in a 12-month running smoothed sunspot number of 59.9 centered on May 2013. Following the curve of the 12-month running smoothed values, a smoothed sunspot level of 79 is expected for February 2014, plus or minus 14 points.

The geomagnetic activity as measured by the planetary A-index (Ap) for November 2013 is 5. The 12-month smoothed Ap index centered on May 2013 is a steady 7.0. Geomagnetic activity should be much the same as we have had during January. Refer to the Last-Minute Forecast for the outlook on what days that this might occur (remember that you can get an up-to-the-day Last-Minute Forecast at <http://SunSpotWatch.com> on the main page).

I welcome your thoughts, questions, and experiences regarding this fascinating science of propagation. You may e-mail me, write me a letter, or catch me on the HF amateur bands. Please come and participate in my online propagation discussion forum at <http://forums.hfradio.org/>. If you are on Facebook, check out <http://www.facebook.com/spacewx.hfradio> and <http://www.facebook.com/NW7US>. Speaking of Facebook, check out the CQ Amateur Radio Magazine fan page at <http://www.facebook.com/CQMag>.

I'll be keeping my ears to the radio, hoping to hear you on the air. Happy DX!73, Tomas, NW7US

Oops…

It's all Greek, um, Russian, to us…

Reader Doug Bailey, K0FO, spotted a few errors—mostly in syntax—in our November article on Russian CW. There were a couple of typos in the table of compass directions shown in the original fig. 4 that resulted from "translating" the Cyrillic characters into a form that our computers could read. Corrections are below.

In the original fig. 11, Doug says, *"You translate the phrase contained in the box as 'No please don't reply in Russian at 40 WPM!' It doesn't say that at all. It tries to say, 'Sorry, this is all the Russian that I know.' The Russian word beginning the phrase means, colloquially, 'Sorry' or 'Regrets,' but is more commonly used in common speech when referring to the action of another person as in 'Sorry you're a dimwit.' In the context of this phrase, a Russian would more likely use a word which means 'unfortunately'."*

The "please don't reply in Russian at 40 WPM" line in the caption was an (obviously poor) editorial attempt at humor. The correct English is in the main text. Regardless, a modified form of "Unfortunately, this is all the Russian I know" is provided here as well. Sorry that we're dimwits!

North	СЕВЕР
Sent as	S E W E R
South	ЮГ
Sent as	ĪM G
East	ВОСТОК
Sent as	W O S T O K
West	ЗАПАД
Sent as	Z A P A D
Southeast	ЮГО- ВОСТОК
Sent as	IM G O W O S T O K
Southwest	ЮГО - ЗАПАД
Sent as	IM G O Z A P A D
Northwest	СЕВЕРО - ЗАПАД
Sent as	S E W E R O Z A P A D
Northeast	СЕВЕРО- ВОСТОК
Sent as	S E W E R O W O S T O

Revised fig. 4, showing corrected spellings of compass directions.

Clarification

In our "On the Cover" column in the November issue, we mentioned that the radio-computer interface used by "cover model" Chris Yody, KE7JBF, was designed and built by the Baton Rouge Amateur Radio Club in Louisiana. We received the following clarification from Dana Browne, AD5VC, one of the interface's designers:

Two members of the Baton Rouge Amateur Radio Club designed the digital interface mentioned in the article. I am one of them. However, we do not build the interfaces for other clubs, despite what the article implies in the last paragraph.

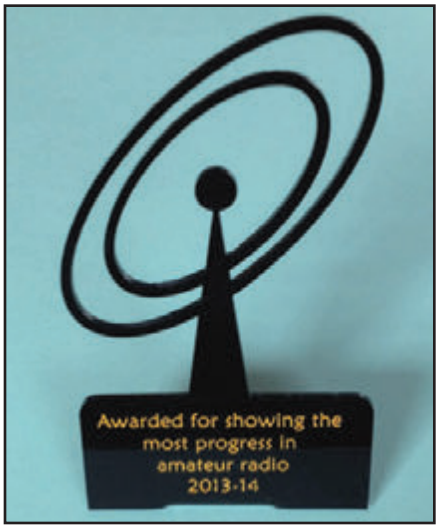
We designed it as a kit building project for our club members. Because of the minimum order needed for the circuit boards, we ordered extra boards and produced unbuilt kits that the club sold as a money-raiser. In fact, the Seattle club that you highlighted bought a large number of kits and then built them themselves.

We are now getting inquiries from hams about having us build interfaces for them, and we have no plans to do that. We may in the future do another kit run, but there is not enough interest yet to make it profitable to the club.

We apologize for any confusion.

(disclaimer)	ЖАЛЪ, ЗТО - ВСЕЬ РУССКИЙ, ШТО Я ЗНАЮ
sent as:	V A L X ŪĪ T O W S E X R U S S K I J Y T O ĀĀ ŪĪ N A ĪM

Revised fig. 11, no longer suggesting that anyone (except the editor) is a dimwit!



Club Trophies from SOTAbeams

Richard Newstead, G3CWI, SOTAbeams owner, says that "one of the most common problems that radio club members face is retaining and nurturing the interests of newer licencees. One way in which this might be done is to introduce some club awards. These could be for showing the most progress in the hobby, working the most DX or even for attending the most meetings."

To help clubs in that effort, SOTAbeams has introduced a club trophy that can be custom-engraved The trophy is laser cut from black gloss acrylic and is priced at £4.99 + p&p in the UK, or just over $8US (plus shipping) at current exchange rates. For more information, visit <http://www.sotabeams.co.uk/club-trophy/>.

SpeeCup Bluetooth™ Speaker-Mic

If you have a Bluetooth™-enabled ham rig, you might want to check out the SpeeCup hands-free speaker-mic. Designed to fit in a standard automotive cup holder (thus the name), the SpeeCup will also fit in many bicycle water bottle holders or can just stand on its own. The upward-facing speaker also features "gesture control," which allows you to answer phone calls or change music tracks with a sweep of your hand over the top of the speaker. *(Whether this would work as a PTT control for a Bluetooth-equipped ham rig is probably a function of the rig's Bluetooth circuitry—ed.)* It also has an audio line out jack for connecting to a home stereo, etc., and has been designated as a 2014 honoree in the Consumer Electronics Show Innovations Design and Engineering Awards. The SpeeCup retails for $129.95 and is available directly from the manufacturer at <http://speecup.com>.

The SpeeCup Bluetooth speaker-mic is designed to fit in a car's cup holder or a bicycle water bottle holder.

RF Torque Wrenches from Pasternack

We must admit that our first mental image here was of a wrench designed to allow a precise amount of RF to pass through a connection, but then realized that in critical applications, it is important for RF feedlines to neither be too loose nor too tight, and that's what these wrenches are about.

The PE5019-XX series of break-over torque wrenches allow accurate mating force (measured in "inch-pounds") when assembling and disassembling a variety of RF coaxial connectors. The unique feature of break-over torque wrenches, according to the company, is that they will "break" at the pivot point once the torque value force is achieved, effectively preventing the user from over tightening the coupling nut on the connector. A broad range of break-over torque wrench sizes is available. For more information, visit <http://bit.ly/19SVsxR>.

Two New Product Lines from RFMW

Sunny Electronics *FastClock™* Oscillators: Circuit designers and experimenters may want to check out the new line of Sunny Electronics *FastClock™* crystal and voltage-controlled oscillators, now offered by RFMW, Ltd. The devices offer a wide choice of frequency, voltage and output combinations, and operate from 10–1500 MHz. For more information, visit <www.rfmw.com>.

Florida RF Labs "Lab Flex" Cable Assemblies: RFMW is also supplying and supporting six different types of cable assemblies from Florida RF Labs. The Lab-Flex® assemblies offer >90 dB of shielding effectiveness and a wide range of connectors including, but not limited to, 7/16, TNC, SMA, 2.4mm, 2.92mm and 1.86mm interfaces in various configurations. Standard lengths are 24, 36, 48 and 72 inches. The cables support frequencies up to 65 GHz. For more information, visit <http://bit.ly/GX7C0B> or call RFMW at (877) FOR-RFMW.

Florida RF Labs "Lab-Flex" cables, available from RFMW. (Courtesy RFMW, Ltd.)

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For YAESU-Vertex VX-5R/s, VX-6R, VX-7R/b, VX-7Rb, VXA-700:
FNB-80Li Li-ION battery 7.4v 1600mAh $44.95
E-DC-5BA DC Power & Charge cord (NEW) $19.95
NC-72BA AC-DC Power / Battery Charger $17.95

For YAESU-Vertex FT-60R,250,270R; VX-110,120,150,170,177,180,210
FNB-83xe eneloop 7.2v 2100mAh $49.95

For YAESU-Vertex FT-817 (PRE-CHARGED); (E-DC-5BA DC cord $19.95)
FNB-72xe eneloop 9.6v 2100mAh $49.95

For YAESU-Vertex VX-1R : (RARE; has custom-designed PCB)
FNB-52Li Li-ION battery 3.7v 750mAh $29.95

For YAESU-Vertex FT-50R,40R,10R; VXA-100: (E-DC-5BA: $19.95)
FNB-41xs Hi-Watt battery 9.6v 1450mAh $52.95

For YAESU FT-11R, FT- 41R, FT-51R, etc. (HIGH POWER battery):
FNB-38h Hi-Watt battery 9.6v 750mAh $39.95

For YAESU FT-530,76,26,416,415,816: (E-DC-5BA: DC Pwr cord $9.95)
FNB-25x Ni-MH battery 7.2v 1200mAh $32.95
FBA-12 6-cell AA Battery Case $22.95
FNB-27xs Hi-Watt battery 12.0v 1450mAh $49.95

For YAESU FT-411, 470, 73R, 33R, 23R etc: (WC-12 wall charger $12.95)
FNB-12xh Ni-MH batt. 12v 1250mAh $39.95
FBA-17 6-cell AA Battery Case $19.95

For ICOM IC-92AD (D-STAR): (CP-11L: DC Pwr/Chg cord $19.95)
BP-256 Hi-Watt Li-ION batt. 7.4v 1620mAh $44.95

For ICOM IC- T70A/E; IC-V80A/E/SPORT, F3003, F4003, etc:
BP-265L Li-ION batt. 7.2v 2200mAh $46.95

For ICOM IC-T90A/E; IC-91A, IC-91AD, IC-80AD (D-STAR), etc:
BP-217 5W Li-ION battery 7.4v 1600mAh $44.95
CP-11L DC Power & Charge Cord (fits IC-92AD too) $22.95

For ICOM IC-V8,V82, U82, F3, F4GS/GT, F30,40GS/GT,A24,A6, etc
BP-210N Hi-Watt battery 7.2v 2000mAh $44.95

For ICOM IC-T8A/E/HP; T81A/E; A23,A5: (WC-AIC Wall Chrgr $12.95)
BP-200XL Hi-Watt battery 9.6v 1450mAh $59.95
BP-197h 6-cell AA Battery case (Hi-Watt) $29.95

For ICOM IC-W32A/E, T7A/E, T7H, Z1A/E, T22A, T42A, W31A/E :
BP-173x Hi-Watt battery 9.6v 1450mAh $59.95
BP-170L 6-cell AA Battery case (Hi-Watt) $25.95

For ICOM IC-2/3/4SAT, W2A, 24AT, 2/4SRA, R1: (BC-105A: $22.95)
BP-83xh Ni-MH battery 7.2v 2200mAh $39.95

For ICOM IC-2/02/03/04AT,2/4GAT etc; Radio Shack HTX-202/404 :
IC-8 8-cell AA battery case (w/ Charge Jack) $24.95
BP-202e Eneloop- Rad. Sh. 7.2v 2100mAh $34.95

For KENWOOD TH-D72A/E: (CP-KE12: DC Pwr & Chg cord: $19.95)
PB-45L Li-ION batt (NEW) 7.4v 2000mAh $44.95

For KENWOOD TH-F6A, TH-F6E, TH-F7: (CP-42L- DC cord: $9.95)
PB-42L Li-ION battery 7.4v 2000mAh $44.95
PB-42XL Li-ION battery 7.4v 4000mAh $59.95
EMS-42K Drop-in Rapid Charger for PB-42L/XL $49.95

For KENWOOD TH-G71/K, TH-D7A/AG/E: (CP-39: DC Pwr cord $9.95)
PB-39h Hi-Watt Ni-MH batt. 9.6v 1450mAh $54.95
BT-11h 6-cell AA Battery Case (Hi-W) $24.95

For KENWOOD TH-79A/E, 22A/E. 42A/E etc: (CP-79: DC cord $9.95)
PB-34xh Hi-Watt NiMH batt. 9.6v 1200mAh $39.95

For KENWOOD TH-78A/E,48A/E,28A/E,27A/E: (CP-17: DC cord $9.95)
BT-8 6-cell AA Battery Case $14.95
PB-13xh Ni-MH battery 7.2v 1800mAh $39.95

For KENWOOD TH-77A/E,75A/E,55A/E,46AT/E,45AT,26A/E,25A/E:
PB-6x Long Life Ni-MH battery 7.2v 1600mAh $36.95

For KENWOOD TH-205A/E,215A/E,225A,315A: (Wall Charger $12.95)
PB-2 Std. Ni-Cd batt. 8.4v 800mAh $29.95

For KENWOOD TR2500, TR2600: (Wall Charger $12.95)
PB-25-26 Std. Ni-Cd batt. 8.4v 800mAh $29.95

For ALINCO DJ-V5, DJ-V5TH : (CP-46: DC Pwr/Chg Cord $9.95)
EBP-46xh Ni-MH batt. 9.6v 1450mAh $52.95

For ALINCO DJ-195/HP/R,193,196,446,493,496,596: (DC cord $9.95)
EBP-48h Hi-Watt battery 9.6v 2000mAh $44.95

For ALINCO DJ-G5TD/TH/TY; 190T,191T/TD/TH: (DC Pwr Cord $9.95)
EBP-36h Hi-Watt batt. 9.6v 800 mAh $39.95

For ALINCO DJ-580/T, DJ-582, DJ-180/T, DJ-280/T, DJ-480 etc :
EDH-11 6-cell AA Battery Case $22.95
EBP-20x Ni-MH battery 7.2v 2000 mAh $32.95

For ADI AT-600; REALISTIC HTX-204 (Wall Charger is $12.95):
ADI-600x Hi-Watt battery 12.0v 1200mAh $44.95

For STANDARD C228,C528,C558; ADI HT-201, HT-401 etc:
CNB-152xh NiMH batt. 12.0v 1200mAh $45.95
CBP-888 8-cell AA Battery Case (Hi-WATT) $28.95

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BY IRICHARD FISHER,* KI6SN

Capitol Hill and FCC Actions Affecting Communications

Broadcasters and DoD Agree on Sharing 1755-1780 MHz

U.S. House Energy and Commerce Committee Chairman Fred Upton and Communications and Technology Subcommittee Chairman Greg Walden, W7EQI, say that they welcome the National Telecommunications and Information Administration's endorsement of an agreement between the Department of Defense and the National Association of Broadcasters on the relocation of a parcel of government spectrum to shared use.

The agreement was reached after bipartisan committee leadership worked with the Department of Defense, the National Telecommunications and Information Administration, and the Federal Communications Commission. It paves the way for the Department of Defense to move systems out of the 1755- to 1780-MHz band by creating a sharing arrangement between it and the broadcast community in the shared use of the Broadcast Auxiliary Service.

This spectrum is used by news organizations to originate material such as breaking news stories from outside of studio facilities. *(**FULL STORY:** <http://1.usa.gov/18yxltt>. – KI6SN.)*

Kansas Radio Amateur Nabbed As FM Broadcast Pirate

Glen Rubash, KCØGPV, of Dwight, Kansas, has been hit with a $4,000 Notice of Apparent Liability for operating an unlicensed FM transmitter on 88.3 MHz in the city of Manhattan.

The fine had originally been issued for $15,000 but was reduced after Rubash produced documentation showing his inability to pay.

The FCC said Rubash admitted over the telephone to its agents that he installed and owned the station's radio transmitting equipment. According to *Amateur Radio Newsline* (ARN), "he also demonstrated control over the station by stating that he would refuse to surrender the equipment to the agents from the Kansas City Office if required to do so."

The FCC said that "even though Rubash admitted via telephone interview to making the admissions, he later asserted that his statements were based on incorrect information ... In his written response he stated he owned and installed a low-power FM radio transmitter but that it operated within Part 15 unlicensed limits."

In denying most of Rubash's requests, the FCC said it affirmed the NAL, finding he violated Section 301 of the Communications Act by using equipment without the required Commission authorization.

More Musical Chairs at the FCC

Communications attorney Roger Sherman is the new Acting Chief of the FCC's Wireless Telecommunications Bureau, which oversees amateur radio as well as other two-way radio services. Sherman replaces former Bureau Chief Ruth Milkman, who was named Chief of Staff to new FCC Chairman Tom Wheeler.

According to the ARRL, Sherman previously served as the Democratic Chief Counsel to the House Committee on Energy and Commerce, along with other Congressional staff positions.

Oklahoma CBer's Linear Amplifier Brings $15,000 Fine

The FCC has issued a $15,000 Notice of Apparent Liability for Forfeiture to Carlton Lewis, of Enid, Oklahoma, alleging he operated a Citizens Band radio with an external power amplifier in violation of Part 95 rules.

In May 2013, agents from the Enforcement Bureau's Dallas office tracked down a strong signal on 27.1850 MHz — CB Channel 19 — coming from Lewis' residence.

For more than 30 minutes, no one answered an FCC agent's knock at the door. Ultimately, Lewis appeared and showed the agent his CB transmitter, which was warm to the touch.

According to an ARN report, "the agent observed that no coaxial cables were connected to the CB transmitter but also noted the coaxial cable coming into the residence and traced it to a linear amplifier hidden behind a sofa. The linear amplifier was also warm to the touch. Lewis did not respond when asked whether he had used the linear amplifier."

Prior to his most recent citation, the FCC noted Lewis' CB station had been issued two warnings from the Dallas Office, advising him that an amplifier voided his authority to operate. Based on previous experience, the FCC ruled a $15,000 fine was warranted.

Commission Upholds Fine for Marketing Unauthorized RF Devices

The FCC has issued a monetary forfeiture in the amount of $14,000 to Custom Interface Technologies, a Division of Thornstar Corporation, in Joshua Tree, California for "willfully and repeatedly violating rules against manufacturing and marketing of unauthorized radio frequency devices in the United States."

The company, alleged to have marketed uncertified video assist transmitters, told the FCC it did not have the ability to pay the $14,000 fine. "However in affirming the forfeiture amount the FCC says while Custom Interface Technologies did provide the Commission with three years of tax returns and a bank statement to support its claim of an inability to pay, after review of these financial documents that the FCC declined to reduce the forfeiture amount and that "the $14,000 fine is warranted."

This Month's Contributors

Sources for information in this month's Washington Beat include the U.S. House Energy and Commerce Committee, Amateur Radio Newsline, CQ Newsroom, ARRL, and other published reports. – KI6SN

*c/o CQ magazine

Photo A. *Syracuse University's Lubin House in midtown Manhattan was the site of September's Audio Industry Summit on the future of radio. (Photography courtesy of W2VU)*

There is good news for broadcast band DXers who like to listen to more than just a station's callsign and city at the top of the hour. Top radio industry execs say 'live and local' are making a comeback …

CQ Plus Reports: 'The Power of the Microphone'

BY RICH MOSESON, *W2VU

Broadcast-band DXing was my first love in radio. In addition to the thrill of tuning in faraway stations, I was captivated by hearing about what was going on in those other places – the local news, the local commercials, the challenge of trying to figure out where the station was before hearing the ID at the top or the bottom of the hour.

The voices coming through the speaker into my bedroom drew me in . . . into radio as a hobby and into a career in radio, television and print journalism. It all started with those voices.

In recent years, though, the uniqueness of local radio has in many places given way to homogenized programming and shows delivered by satellite from New York or some other big city. Now, satellite radio is even taking local radio stations out of the equation, along with Internet "radio" sources such as *Pandora* and *iHeartRadio,* which lets listeners eliminate those voices altogether and simply package music that is sent over the Internet to your smartphone or even your car.

The Tide is Turning

Good news. "Terrestrial radio" is fighting back. And it is sending to the front lines the DJs and news reporters who

Photo B. *Newhouse School of Public Communications Dean Lorraine Branham said her advisors told her to call the session an "audio" industry summit because "radio" was too limiting. "But it's all about radio," she said in opening remarks.*

made broadcast radio such an important part of so many American communities, but who, in recent years have frequently been marginalized.

"Radio is not about music," CBS Radio President Dan Mason told an Audio Industry Summit in late September. "Radio is not about news, or traffic or sports. Radio is about the power of the microphone."

His feelings were echoed by Bob Boilen, host of "All Songs Considered" on National Public Radio. "I want somebody to talk to me," he said. "I want a storyteller. That's what I want out of the medium ... That's where radio wins."

The summit was organized by Syracuse University's Newhouse School of Public Communications and held at Lubin House, the university's "outpost" in New York City[1], **Photo A.**

Newhouse Dean Lorraine Branham, **Photo B,** opened the event by discussing her memories of growing up listening to the radio in Philadelphia, and pointed out that the meeting was being called an "audio industry summit" because "my advisers told me 'radio' was too limiting." Then she paused and added, "But it's really all about radio."

About 100 people attended[2], including two groups of panelists. Most were top executives of various media companies and, in a reflection of where we continue to be as a society, the group was overwhelmingly male, white, and over 50.

The first panel, **Photo C,** was supposed to discuss the current state of the radio industry and the second, **Photo D,** was supposed to look at its future, particularly in the face of competition from satellite and Internet radio "disruptors" such as Pandora and iHeartRadio (both of which were represented on the second panel). But they kind of got all that mixed up and both groups talked about both the present and the future.

Live and Local

The first panel was made up of executives of radio stations or station groups, including CBS Radio's Dan Mason, **Photo E**; Patrick Walsh, CEO, COO and director of Emmis Communications, which is one of 10 biggest radio station owners in the country; Deon Levingston, Vice President and General Manager of WBLS in New York City, **Photo F**; Buzz Knight, in charge of programming development for Greater Media and former Program Director of various stations, including the iconic WNEW-FM in New York; and Ed Levine,

Photo C. *The first of two panels was made up of top executives of traditional radio stations or station groups. (See text for a "Who's Who")*

Powerhouse WABC Radio Makes Shift to 'Local'

New York City's WABC-AM 770 is making dramatic changes to its talk radio line up, with a greater emphasis on "local" hosts and issues. Details:< http://www.CQPlusDigital.blogspot.com >.

Photo D. *The second panel, on radio's future, included a mix of people from "terrestrial" radio and Internet-based radio services.*

President and CEO of Galaxy Communications, which owns 14 radio stations in upstate New York.

All of the speakers were very positive and upbeat about the state of the radio industry and its future, with an emphasis on the need for radio stations to return to being "live and local" in order to set them apart from satellite and Internet competition. "Local is our job," said Levingston. Knight said his company is "now seeing the importance of good content and good personalities," citing the midday DJ at his company's station in Philadelphia – Pierre Robert – as "the heart and soul of Philadelphia."

Walsh echoed the theme of bringing the "local" back to local radio, while acknowledging that the two largest radio station groups in the country (Clear Channel and Cumulus) do not yet share that vision. But as far as his company is concerned, Walsh said, "Strategically, local makes sense, it's where our growth is. Our personalities drive us." And he added, "Being local is what differentiates us, what makes us special and different ... that connection with the listeners."

This attitude should come as welcome news to broadcast-band DXers, who should once again be able to reap the benefits of listening to uniquely local programming as they spin the dial, seeking out new stations in faraway places.

Photo E. *CBS Radio President Dan Mason told the group that radio is not about music, news or sports, but rather about "the power of the microphone."*

Photo F. *Deon Levingston, Vice President and General Manager of WBLS in New York City, noted that "local is our job."*

Goin' Mobile ...

A few years back, I wrote a guest editorial in *Popular Communications* about missing my transistor radio and lament-

ing that I couldn't listen to the baseball playoffs while shopping at my local supermarket[3]. Well, that topic also came up, with Walsh pointing out that the radio industry has recently put together an "app" that activates an FM receiver on some Android phones offered by Sprint. "On average," he noted, "people look at their phones 156 times a day, or every six minutes." Or as Mason put it, "the cell phone has become the new transistor radio."

WBLS's Levingston welcomed that development, noting that in recent years, "as an industry, we've lost our portability." He pointed out that there were three distinct stages of technological development in broadcast radio, starting with the home console radio that plugged into the wall. The next big leap was into cars in the 1930s and '40s, followed by the introduction of the battery-powered transistor radio in the '50s and '60s. But the last portable radio with any popularity, he pointed out, was the Sony Walkman in the 1980s. "This app," he said, "will help us regain our portability."

Car radios are now doing more than just playing music, and that was part of the focus of the second panel, which included representatives of the "disruptors" as well as traditional broadcasters. Along with "mobile devices" (smartphones, for example) getting radios, car radios are beginning to provide for Internet information sources, such as Pandora or iHeartRadio. And, just as the frequency display on your scanner or ham rig has now become a multifunction display providing a variety of information, so are car radio displays now offering station, song, and artist readouts as well as album art and other visual information.

Photo G. *Kevin Straley of TuneIn discussed the fact that after the Boston Marathon bombing, his service's listeners "tuned in" to Boston police scanners as much as they did to local and national news sources.*

Speakers on the second panel included Chris Oliviero, Executive Vice President of Programming for CBS Radio; NPR's Bob Boilen (introduced earlier); Owen Grover, vice president of "content partnerships" for Clear Channel Entertainment and a co-developer of iHeartRadio; Tommy Page, vice president of artist and brand partnerships at Pandora; Kevin Straley, VP of programming at TuneIn; and Ryan Delaney, a news reporter for public radio station WRVO, a 2010 Newhouse School graduate and token "young person" on the panels. The moderator was Doug Quin, a professor at the Newhouse School and do-director of its audio arts department.

Grover, Page, and Straley emphasized they consider the newer, Internet-based, entertainment media to be *additive* to traditional terrestrial radio rather than competitors, providing listener-controlled programming.

"I believe Pandora is the future," said Page. "The future is streaming, and small artists. Multiple platforms encourage new and small artists." He noted that the iPhone had prompted exponential growth in Pandora. "Two years ago, 80 percent of our listeners were on desktops. That's now reversed, but with no loss of the desktop audience. It has all been growth on the mobile side."

Grover, however, pointed out that radio over IP is still very young, reaching only 50 percent of the U.S. population, while streaming video already reaches 90 percent. "There is a lot of room for growth," he noted.

Not Just Broadcast

One aspect of the growth in radio over IP that should be of interest to many *CQ Plus* readers is that the medium is not restricted to music or even to broadcast radio stations. TuneIn's Kevin Straley, **Photo G,** said that in the aftermath of last spring's Boston Marathon bombing, "TuneIn had links with WBZ news radio, WRKO talk radio, CNN for a national perspective, and Boston police scanners . . . which rated as high in listenership as the national media sources."

Straley later said that "radio is just one part" of what TuneIn offers," noting that "we have more than 70,000 live streams, including scanners and FAA airport towers."

Indeed, when clicking the "local" tab on TuneIn.com's website from my home computer in northern New Jersey, the menu of choices included not only all of the FM, AM, and HD broadcast stations in the metro New York area, but a wide variety of amateur radio repeaters and public safety agencies as well.

Personal Connections

The overall emphasis of all the speakers – whether representing traditional broadcasters or Internet-based sources – was the necessity of engaging individual listeners. The Internet radio people see that as coming through providing listeners with individual control of what they hear, while the broadcasters are focusing on the personal connections and community involvement that come with the return of live, local disk jockeys.

"We want the DJ to once again be the biggest star in town," said CBS's Olivieri.

"Radio stations have always had the reputation of calling people to action," noted his boss, Dan Mason, while Clear Channel's Owen Grover later echoed that thought, saying, "My colleagues truly plant beacons in the center of tribes." But, these radio industry leaders are learning, those beacons are only jukeboxes if they don't use "the power of the microphone."

*(**LISTEN:** To a 2-hour, 45+ minute podcast of the Audio Industry Summit at <http://bit.ly/16HcJOE>. – W2VU.)*

Notes:

1 - The city of Syracuse is 250 miles from New York City, yet many of Syracuse University's students come from and return to the New York City area. Since 1965, the university has operated Lubin House as its "remote base" in New York City, featuring events for alumni, current, and prospective students. To learn more, visit <http://lubinhouse.syr.edu/>.

2 - The summit was also streamed live on SU's student radio station, WERW, making it accessible in real time to Newhouse students at the main campus in Syracuse. A podcast was also posted for later listening at <*http://bit.ly/16HcJOE*>.

3 - *Popular Communications*, August 2010

Digging Into Specifics of the Commission's 'Revitalization' Proposals

The FCC Moves to 'Fix' AM Broadcasting

BY ROB de SANTOS, K8RKD

This extended version of *Horizons* explains the recent FCC Notice of Proposed Rulemaking (NPRM) regarding the U.S. AM broadcast band and what it might mean for medium-wave broadcasting in the United States.

We've been covering the changes affecting the AM broadcast band for well over a year now.[1]

An AM Haze and Malaise

It's no surprise to many of us that the AM band is suffering badly from neglect and declining numbers of listeners, **Photo A**. After nearly 18 months and on the doorstep of a new leader at the FCC, a NPRM was issued on October 31, 2013 laying out the plans for the future. *(IN DEPTH: See FCC NPRM 13-139: "Revitalization of the AM Radio Service" at <http://fcc.us/18UO2xm>, Photo B. – K8RKD.)*

The FCC has proposed to make a number of changes, some that will be popular, and others likely to be controversial even among station owners and managers. It has not opted, at least for now, to go along with some of the more radical proposals including moving AM broadcasters to an expanded FM band or taking the band to entirely digital. With these choices off the table, it remains to be seen if what is now proposed will have the desired effects on the AM band.

As I've noted in earlier columns, listeners to medium-wave or AM broadcasting in the U.S. have declined radically over recent decades. As late as the mid-1970s the majority of all radio broadcast listening in the United States, and indeed worldwide, was in the AM mode — not FM. Today, by some measures, the band garners less than 10 percent of all listening hours in the U.S. and with the advent of digital services, that decline is accelerating. Among the youngest demographic sectors, the band is almost completely ignored. In the NPRM, the FCC noted that only 4 percent of listening time among those aged 25-34 was to AM.

Among the other issues facing AM, these include increased interference and low fidelity. The interference increase is due to several factors including the proliferation of more and more electronic and computerized devices, slack enforcement of the FCC's own standards for interference, the overlay of digital services on the band, and the changes in the FCC rules allowing higher power and extended hours for many stations.

It's No Longer the 1930s

What makes AM effective as a means to reach wide geographic areas

Photo A. *FCC proposals may bring big changes to AM radio across the United States. This Victor Model 3-RF-91 AM-FM radio was found by a lucky looker at a yard sale. After a bit of tender-loving care, it works and looks great. "This thing has an 8-inch speaker and sounds fantastic," writes the photographer. It may be hearing a whole new type of medium wave if all the FCC proposals are adopted. (Courtesy of AlexKerhead via Wikimedia Commons)*

email: <commhorizons@gmail.com>
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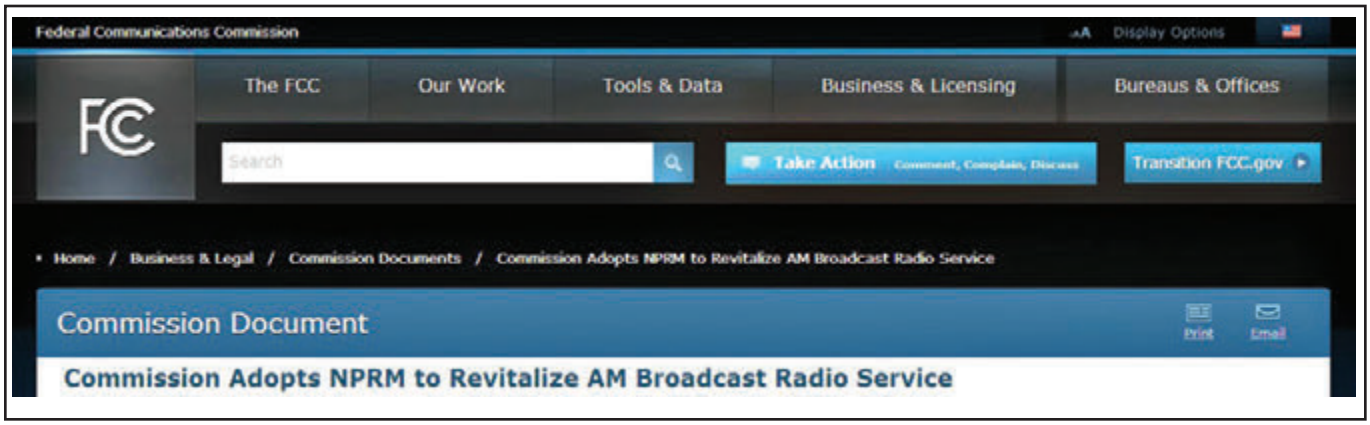

Photo B. *The full text of FCC Notice of Proposed Rulemaking (NPRM) 13-139, regarding U.S. AM radio, can be found at "Revitalization of the AM Radio Service" at <http://fcc.us/18UO2xm>. It can be viewed as a Word, PDF, or text document. (Internet screen grab)*

at night also makes it subject to nighttime interference. The low solar cycles haven't helped either as skywave propagation at night has suffered. While DXers and radio lovers might not be happy with the past changes to the AM band, the country is a different place than it was in the 1930s when the basics of the AM band plan were first fixed and "clear channels" set by regulation and treaty.

It was a reasonable assumption in 1934 that having a few stations that could be heard by most of the country provided a guaranteed communication backbone for the U.S. Many local stations back then could not have afforded to be on the air well into the night even if they were permitted to do so. Those assumptions may no longer be true.

Judging from the programming on many of the remaining stations with "clear channel" frequencies, the station programmers no longer think it important to cater to 30-or-more states at night and only to their local audiences even if they occasionally acknowledge others are hearing them.

Changing Technologies

With satellite and fiber distribution of programming, listeners in remote areas can be reached in emergencies without necessarily needing clear-channel stations. State emergency radio networks now reach every geographic corner of their territory in many cases. The advent of the Internet has made it possible for very low-wattage stations to be heard worldwide.

Changes in music listening and the growth of talk radio have also had an impact. FM brought static-free sound to audiences and imaginative programming saw migrations of listeners begin to move away from AM in the 1970s.

The number of AM stations in the U.S. has declined by an average of nearly 10 per year since 1990 while the number of FM stations has doubled. There are now more than three times as many FMs as AMs. Despite actions by the FCC in 1990-91, 2005, 2006, 2008, 2009, and 2011, the tide has not turned. The emergence of highly popular syndicated talk radio in the mid-1980s briefly seemed to save the band but now those very talk formats are increasingly moving to FM.

Specifics: What the FCC is Proposing

The NPRM contains at least six major changes. Some of them are highly technical adjustments that matter mostly to stations' engineers and owners and others have more broad impact. I'll try to give you a simplified explanation of each and what it means to hobbyists and listeners.

FM Translator Expansion

The agency has proposed to open an exclusive window for AM station owners to apply for new FM translators. This window would be open after the FCC has concluded processing the recent applications for low-power FMs for community and non-profit applicants and disposed of the thousands of remaining applications for FM translators from current FM station owners. ***(NOTE:*** *Some of the latter have been pending for many years. – K8RKD)*

There are some unique aspects to this particular proposal. The most important is that it would be restricted to current station licensees and permit owners and the translator could not be sold except with the primary station license.

The translator would be permanently linked to the primary AM station. In addition, the coverage area of the translator must be the smaller of either the station's primary daytime coverage area (2mV/m contour), or a 25-mile radius of the current transmitter site. ***(IN DEPTH:*** *See the accompanying "What Does mV/m Mean?" – K8RKD.)*

For many lower-power AMs or those with limited or no nighttime service, they would now be able to serve their primary market 24 hours a day.

Reduced Daytime Coverage Standards

One of the issues facing many commercial AM station owners, particularly those in highly urbanized environments, is that making any change to your transmitter site or antenna system that would improve listening for the majority of the audience or the stations market potential is not possible.

If the change reduces coverage of the primary city of license below 80 percent of either the population or area, whichever is higher (5mV/m contour for the techies), it will not be permitted.

The FCC has proposed to change the standard to be similar to that in effect for non-commercial stations: 50 percent of the population or 50 percent of the area the community of license at a minimum of 60 dBm – decibels per microvolt of signal strength.

Photo C. *"The interference-free contour means that many stations on a given frequency have a minimum protected area on that frequency," K8RKD writes. "For example, for clear-channel stations such as 1030 WBZ in Boston this can be half or more of the U.S., Canada, or Mexico." (Internet screen grab <http://cbsloc.al/1cNmZmG>)*

It is possible that any future waivers of the existing standard would also be eliminated. For readers, this would certainly affect the behavior of stations for DXing and casual listening purposes. Many stations would become unlistenable in some local areas while others would immediately become DX opportunities due to the facility improvements.

Station owners would gain considerable flexibility in making transmitter and power changes.

Nighttime Coverage Standards

Many AM stations have different antenna or different transmitter sites in order to meet the different technical requirements for daytime and nighttime service.

What's Does mV/m Mean?

The formula mV/m stands for milliVolt per meter and is a measure of the strength of the electric field at a certain distance.

Simply, if there is a voltage difference of 1 Volt and the two places you are measuring are 1 meter apart that is a strength of 1 V/m.

When talking about radio waves, the electric field varies inversely with distance. This is a one-dimensional measurement and it only varies by the inverse — *not* the square of the inverse. Double the distance and the electric field is half as strong. One hundred times the distance and one-hundredth the radio wave field strength.

The best way to view this as a listener is to think in terms of what your radio is capable of picking up. A typical car radio should be able to pick up any station which meets the 2mV/m standard for your listening point — local noise levels notwithstanding. – *K8RKD.*

As many readers know, AM propagates as groundwave during the day and skywave at night. Skywave propagation is similar to shortwave propagation while groundwave is limited by distance, geography, and power. *(**IN DEPTH:** In skywave propagation, signals reflect off the ionosphere and can be heard at great distances <http://bit.ly/1fZgToo>. – K8RKD.)*

During the day, stations must meet the aforementioned 80 percent coverage requirement. At night, there is the additional requirement that the signal cannot breach the guaranteed interference free contour of other stations.

The interference-free contour means that many stations on a given frequency have a minimum protected area on that frequency. For example, for clear-channel stations such as 1030 WBZ in Boston, **Photo C**, this can be half or more of the U.S., Canada, or Mexico. As such, other stations are thereby limited to their local market and nothing else. As you might guess, with skywave propagation, that is a tall order since the signal might bounce off the ionosphere and be heard several states away.

The FCC has proposed to modify the 80 percent rule. The FCC is proposing that stations be required to serve only 50 percent of their area or population, whichever is higher, (nighttime contour again) but still be subject to the non-interference rule.

Eliminate the 'Ratchet Rule'

The 'Ratchet Rule' regards a more technical requirement which affects higher power AM stations — in FCC parlance, Class A and B stations. The current rule requires AM stations to reduce — or *ratchet back* — their power levels such that it reduces skywave interference to other stations on the same frequency if they apply to make any facility (transmitter, power, antenna, etc.) changes to their license.

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This rule was part of the FCC changes made back in 1991 when the band expanded adding channels up to 1710 kHz. The intent was to reduce interference. However, the FCC has now concluded it hasn't worked as intended.

Various petitions submitted to the FCC have argued the rule does, in fact, discourage stations from making improvements since any facility change such as transmitter relocation, antenna pattern change, and so on, mean the station must reduce its power to comply with the rule.

The result is a loss of interference-free service at night at minimal benefit to other stations. The stations most disadvantaged by the rule are those which are the oldest on the air while newer stations already accepted the interference conditions when they went on air and thus get little benefit.

The FCC has proposed to delete the rule entirely. For listeners outside the station's primary area of service, this might not have much impact. It won't reduce the interference-free zones of other stations on the same frequency. Their protection is retained. If anything, it will improve the quality of some stations heard at night outside their primary coverage area and have some marginal reduction in DXing of weaker stations on the same channel.

Modulation Dependent Carrier Level Control (MDCL)

MDCL is another very technical change. At present, with only a few dozen exceptions by waiver or experimental permit, broadcasters generally do not vary the power of the carrier frequency or the sideband as modulation is changed.

To do so would result in the input power to the antenna below levels permitted under FCC regulations. The FCC is proposing to permit carrier and sideband power to vary with modulation — generally, full/higher power maintained at low modulation and reduced power at higher modulation.

For station owners this could mean reduced costs due to lower power consumption by the transmitter. The FCC would permit stations to implement MDCL without prior FCC authorization and do so even if they provide hybrid (analog and digital) service.

If all of that means nothing to you, don't worry. Since only a few dozen stations have tested this. It is too early to say what it would mean for listeners, if anything. A few critics of poor AM signals would probably welcome a change that reduced modulation as it might improve the sound of some over-modulated stations.

Modified AM Broadcast Antenna Requirements

Currently, the FCC requires that any AM station seeking to make a change to its facilities, transmitter, and so on, must ensure that the new installation still meets the requirements for the class and power levels of their license. This is quite reasonable.

In practice, the station must demonstrate that either the antenna is of a certain minimum height above the surrounding terrain or its effective signal strength meets the minimums over its service area.

The FCC has proposed a 25 percent reduction in the minimum field strength requirements for each class of AM stations, which would provide more relief and flexibility to station owners to make station modifications. For fringe listeners, this might mean less service but it could also mean better service for the primary market if the newer antenna and transmitter system is of higher quality.

What and When

The NPRM requests extensive comment from the radio industry and public and it is quite possible that one or more of these proposals will change significantly before they are implemented by the Commission.

At this writing, the timetable is still a bit fuzzy. But barring extensions, the comment and reply period will have closed in late January 2014. We can then expect that the FCC will take some time to process all of the input and another NPRM or a Final Report and Order may be forthcoming by the middle of 2014.

These proposals largely address technical and financial concerns of station owners, managers, and engineers. They do little to address the fundamental market issues driving the decline of the band.

At best, they buy time for the radio industry to settle on and propose changes that are more radical. Even if interference is reduced and signal quality and reach enhanced, it is hard to see how these changes will result in stabilizing the band and preventing further declines in the longer term.

What is Your Reaction?

Listeners and DXers have quite a bit to consider here, too. It's time to think about what we want from the band going forward. If you have ideas, write to the FCC and your congressional representatives and let them know. Also, let us know.

Which of the proposals should the FCC adopt? Is it even possible to save the band at this point? What does it mean for your DXing? Share with me your reaction. Write to me via the magazine, or drop an email. We'll publish some of your thoughts in a future article. More on the future of communications in *Horizons* next month.

– K8RKD.

1. Horizons' *"There's A Crisis Brewing in U.S. AM Radio," Popular Communications November 2012*, and the debate that followed: *"The Agony: A Slow Death for AM Radio?" Pop'Comm March 2013.*

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BY RANDALL NOON, KCØCCR

For Valentine's Day: Romancing the Key — A CW Love Story

Computer-generated digital modes require a keyboard with 104 buttons or more — most with more than one function — that have to be touch-memorized in order to enter messages competently into a transmitter.

Computer-generated digital modes then require a matching decoding program at the de-modulation end to convert the digital message into clear copy. Compare this to Morse code, requiring only one button and a trained ear. *Simple.*

When band conditions are poor, a message sent in CW can get through when the same message sent by single-side band (SSB) phone, amplitude modulation (AM) phone, frequency modulation (FM) phone, or RTTY cannot. To match the good signal-to-noise ratio of a modest 5-watt CW signal, an SSB phone signal needs more than 100 watts. *Practical.*

Morse Code also has an inherent rhythmic, musical quality. To some, it sounds pretty. Like a songbird. There are:

- Triplets, such as the letters *S* and *V* and number *3*
- Syncopations, like the letters *C, L, F,* and *Q*
- Short, rhythmic drum patterns, such as the comma, question mark, the letter *H* and the number *5*

In other words: *Musical.*

She: Betty Broome, VE3ZBB

Allow me to add one more Morse code descriptor to the list: *Romantic.*

The explanation of this adjective begins with FISTS member No. 1265, Elizabeth Broome — otherwise known on the air as VE3ZBB.

(**BACKGROUND:** VE3ZBB was instrumental in providing background to this story, along with *The Canadian Amateur*, which carried her story in 1998/ – KCØCCR)

FISTS is another name for the International Morse Preservation Society with more than 16,000 members worldwide. Elizabeth, or Betty as friends call her, earned her Canadian ham license in March 1988. She was the first and only one in her family to hold a license. In 1993, Betty became a member of FISTS.

As many do, Betty started using a hand key, then switched to using paddles, and now uses either one. She enjoys working CW on all the bands, but favors 30 meters, the only all non-phone band, and is not on voice at any time on HF.

When possible, Betty likes to work CW newcomers and encourage them by sending at a word-per-minute speed comfortable for them, no matter how slow. Like the FISTS motto: *when you work VE3ZBB, you have worked a friend.*

He: Rod Newkirk, W9BRD

On August 28, 1991 at 19:58 hours UTC, according to her log, VE3ZBB worked W9BRD on 30 meters at 10.107 MHz. Recalling that QSO, Elizabeth noted that, "for some reason, I knew at that moment that this radio operator was special."

The call W9BRD belonged to Rod Newkirk, an extra-class operator who lived in Chicago. As Betty recalls, "when Rod sent code it was beautiful to listen to and to copy."

Remembering Their Chance CW Meeting

Betty recalls that at first they exchanged signal reports and typical information. Despite poor band conditions that day, they drifted naturally into a lively CW conversation. Rod told Betty that he had been retired for four years and he was, "reliving his teen years with no responsibilities at all, *so you better watch out, lady."*

He said he'd been a widower for 17 years and mentioned to Betty he "... no longer signed any contracts." Rod was enjoying the best retirement hobby of all, he said. Unfortunately, the band kept fading. Finally and reluctantly, VE3ZBB and

Betty Broome, VE3ZBB, and Rod Newkirk, W9BRD, visit Glencoe Beach during their Chicago "first meeting" in 1993.

e-mail: <rknoon@nppd.com>

Mr. and Mrs. Newkirk, married June 27, 1997.

W9BRD had to say *73* (best regards) and end the contact after struggling to keep the conversation going with an uncooperative 10.1-MHz band. Betty hoped they would meet again on the bands. Over a half-hour had passed during their enjoyable CW banter.

As is often said, *and is true*, the final courtesy of a QSO is the exchanging of QSL cards. Betty and Rod were nothing but courteous. They also exchanged pictures. Rod sent along a letter to explain why his QSL had been delayed. It was sufficiently entertaining that Betty shared it with her husband, son, and daughter-in-law.

Rod Newkirk at his and Betty's Yaesu FT-1000. Note Betty's bulletin board with photographs above the transceiver.

Rod's photograph was put on Betty's bulletin board above her transceiver with many other photographs that other operators had sent to her over the years. She listened for his callsign to pop up again on the bands, but did not hear it.

Comfort in Loss via Amateur Radio

In March 1992, Betty lost her husband of 42 years. Adjusting to a new and different life is difficult after 4+ decades. Betty found free time hard to come by. However, amateur radio was a comfortable place where Betty could spare a few minutes now and then with both old and new on-the-air friends. She said it was, "an escape into another world, and an extremely pleasant one."

Almost a Year Later, Love is in the Air(waves)

On July 30, 1992 at 21:25 UTC a little CW magic happened. On 30 meters (10.107 MHz) Betty sent *CQ.* She again heard W9BRD's excellent Morse answer — big surprise! She replied right away. As Betty recalls, "Here again was the person I'd had such an enjoyable CW conversation with 11 months ago."

During the subsequent QSO chat, Rod told her that from time to time he had heard her on the bands, but that she had been busy with another operator. Betty shared with Rod her recent loss.

Since Rod had been through a similar rough patch, he well understood what was going on in her life. The reunion QSO lasted 45 minutes. This time, though, just before the 30-meter band dropped out, they agreed to a weekly schedule on Thursdays.

On-Air Date: Every Thursday ... and Monday

Betty recalls that she looked forward to the Thursday QSOs. Later, a Monday schedule time was added because, as Rod said, "Thursdays only weren't enough."

Soon, they began exchanging letters and telephone calls. In her first letter, Betty mentioned to Rod that he should have been a writer because she enjoyed his letters so much. Rod must have had a fun time telling Betty, who had not been a *QST* reader, that he had been a writer for the American Radio Relay League — having written *QST's* "How's DX" for more than 30 years.

Betty discovered, among many things, that Rod had been a ham since 1937. In earlier years, he had been a contester

Rod sends code on Betty's Heathkit HW-9 QRP transceiver at Green Lake Cottage.

VE3ZBB and VA3ZBB — t'was a wonderful life ...

and had also attained DXCC. Of the many certificates, Rod had achieved 55 wpm CW proficiency from the ARRL. He used a Vibroplex or Bencher paddle for CW, and sometimes used a straight key.

He still liked contesting and knew where to look for exotic stations. Often he was the first to answer their CQ, and then came the pile up. In his later years, however, he gave up contesting and simply enjoyed spending time on the bands making conversation and snagging a DX station or two.

With the increased frequency in CW QSOs, letters, and telephone calls, it was obvious to just about anybody — but perhaps *them* — that Betty and Rod had moved out of the narrow, friendship sub-band into a wider mode. Both Betty's and Rod's families encouraged them to meet.

They Meet in Chicago; Marry in *Shangri-La*

Betty traveled to Chicago in the spring of 1993 and the two CW aficionados finally met — one and a half years after that first CW QSO.

Rod and Betty became engaged in November 1994. Rod became FISTS member No. 7672 on June 1995, and the couple was subsequently married on June 27, 1997 in Ottawa, Canada. Rod was 75 and Betty, 68. *Rod and Betty (Broome) Newkirk*.

Rod subsequently moved to Ottawa, parked his keys and keyers in front of their shared transceiver, and christened their Ottawa cottage *Shangri-La.*

Betty recalls Rod could send code on any type of key, and even made one of his own on a simple wooden board. He was interested in indoor antennas and tested his new ideas in their radio room.

"Our radio room always had some kind of structure set up in it," Betty said, "and he tested each antenna individually, kept records of his designs along with construction details and eventually the contacts he made with them."

Rod became a Canadian citizen, studied and took the Canadian amateur radio examination, and was rewarded with the callsign, VA3ZBB. Sometimes this caused a bit of fun confusion when operators called Rod, Betty, and vice-versa. *They had a wonderful life together.*

Never Ending Stories

Rod continued to write and contributed to the *K9YA Telegraph E-Zine,* at <http://www.k9ya.org>.

Some of the pieces Rod wrote were gathered into *The Rod Newkirk Collection,* which was printed and bound by Lulu <http://www.lulu.com>. It is still available.

As noted by a reviewer, "Rod Newkirk joined the K9YA Telegraph staff in September 2004 with the publication of his story, '*A QSO with CRØOK*.' That first, short article about a legendary Chicago gangster and his predilection for Morse code set the benchmark for the 44 articles to follow."

Rod Newkirk, VA3ZBB (SK)

Sadly, VA3ZBB passed away November 19, 2012 after spending more than three years in the hospital following a stroke.

Betty visited him every day — sometimes two and three times a day. While his left arm was affected by the stroke, his mind was keen and his right hand was still able to operate a key. Betty, however, was not allowed to set up a transceiver in the hospital.

T'was a Wonderful Life

Betty and Rod shared a life that began with a just routine CW QSO. *A wonderful life.*

As Betty noted in *The Canadian Amateur* in February 1998, "Morse code telegraphy is a unique form of communication between two people ... My special way of expressing it would be an invisible thread that ties two people together."

Epilogue: Betty is on the air again and has a new Cushcraft R9 antenna that is working great. Check 30 meters for her CW CQ.

Simple, practical, musical, and even romantic. CW is so cool.

Acknowledgements

Background for this article was gathered from *The Canadian Amateur,* February 1998; page 35.

The Rod Newkirk Collection, by Rod Newkirk, published by the K9YA Telegraph, Robert F. Heytow Memorial Radio Club, is copyright 2011, and contains a story about Betty and Rod. A nice photograph of the two of them is on page 8. The book, by the way, is a fine read. I recommend it.

A personal *thank you* goes to Nancy Kott, WZ8C, FISTS Coordinator who remembers everything and helped kickstart this story. *73, Randall, KCØCCR*

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BY GERRY L. DEXTER*, WPC9GLD

A Big SWL Shuffle for Radio Miami International

Strange twists are occurring in the shortwave world. In the midst of shutdowns, closures, permanent sign offs, penny-pinching governments, and other program producing organizations discontinuing operations, comes an odd situation.

Well-known broadcaster Radio Miami International has shut down its operations from Miami (on 9955). But not to worry, programming will immediately resume from the facilities of the former WYFR. In short, a mammoth turnover has taken place.

WRMI's single-frequency operation has instantly become one of the big boys, with access to Family Radio's 13 high-power transmitters — feeding 23 antennas — providing worldwide coverage.

The "new" WRMI will continue to carry Family Radio's programming for South America and the Caribbean and will add programming from Pan American Broadcasting, normally heard on 15190 from Equatorial Guinea. *Whew!*

That's what you call a "bombshell." Good luck to Jeff White and WRMI. Hopefully he'll have many, many years of successful operations from Okeechobee!

<gdex@wi.rr.com>

REE Relay Station Closed in Latin America

Radio Exterior de Espana has closed its relay station at Cariari, Costa Rica. *Now why would it do that?*

Your guess is as good as mine. REE didn't give us an explanation, but it was probably a money-saving move, although REE *does say* it will try to make up the decrease in coverage by upgrading its main (read that *only*) other facility at Nobeljas, Spain.

Here are three segments of the giant, multi-part QSL Ralph Perry received from the Voice of Han in Taiwan. Top that one for splendor if you can!

The Costa Rican facility has been around since it opened in October 1992.

Pirates, Get Out Your Checkbooks

The Voice of Croatia, Deanovic has been off the air for some time. Now it seems to be serious about it. It is offering its 100-kilowatt transmitter and all of its associated equipment for sale. *Any pirate operators out there with really deep pockets?*

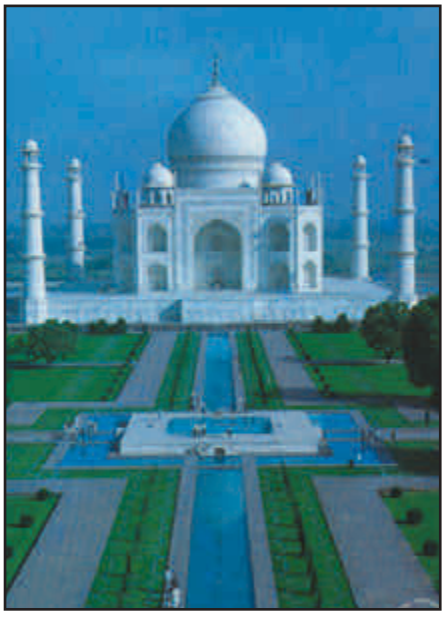
Good News: This Month's Shortwave-lets

• **Radio Aparecida** in Brazil's' like-named city, on 5030, 6135, 9630, and 11855 is installing new transmitters. I presume those will be of the same or higher power than those currently in use, which range in strength from 25 kilowatts down to one.

• I've seen reports to the effect that **Kol Israel** has returned to the air, using 15760 from 1400-1530 Sundays through Thursdays and closing five minutes earlier Friday and Saturday.

• Apparently shortwave from Greece isn't quite dead after all. A new public service broadcaster is in formation. It grew out of the ashes of the now-closed ERT (Voice of Greece). The new service, abbreviated as **NERIT**, initially focused on news reporting and will, presumably, add music and other programming as it develops.

• **Radio Sultanate of Oman** has added Arabic broadcasts on 9540 at 0400-1000.



All India Radio shows the Taj Mahal on this nice QSL received by John Miller in Georgia.

Rich D'Angelo got this nice card from RTE Radio, Ireland, broadcasting special coverage of the All Ireland Hurling Finals via Wooferton.

KVOH — The Voice of Hope confirmed one of its test broadcasts for Rich D'Angelo.

• **The Voice of Russia** says it ceased shortwave broadcasting on January first. There's some question as to whether that report is accurate. It may be that the cessation applies only to its broadcasts in Russian (a.k.a. *Golos Rossii*).

• **Radio Slovakia International** has returned to its former activity of airing half-hour transmissions. This challenging outlet is rarely heard in the early evenings.

SWLers: Now It's Your Turn

Remember, your shortwave broadcast station logs are always welcome. But please be sure to double or triple space between the items, list each logging according to home country and include your last name and state abbreviation after each.

And before you submit them, please proofread them. Several logs this month did not include the time and, thus, had to be omitted. Also needed are spare QSLs or good copies you don't need returned, station schedules, brochures, pennants, station photos, and anything else you think would be of interest.

And please include the transmitter site where appropriate. Also, how about sending a photo of you at your listening post? It's your turn to grace these pages. Feel free to send your photo via email.

Hint: Having your shack photo included would make you the envy of every reader of the *The Listening Post*.

SWL Listings:
For the latest SWL loggings, visit: <http://www.CQPlusListeningPost.blogspot.com>.

world band tuning tips

World News, Commentary, Music, Sports, And Drama At Your Fingertips

This listing is designed to help you hear more shortwave broadcasting stations. The list covers a variety of stations, including international broadcasters beaming programs to North America, others to different parts of the world, as well as local and regional shortwave stations. Many of the transmissions listed here are not in English. Your ability to receive these stations will depend on time of day, time of year, your geographic location, highly variable propagation conditions, and the receiving equipment used.

AA, FF, SS, GG, etc. are abbreviations for languages (Arabic, French, Spanish, German). Times given are in UTC, which is five hours ahead of EST, i.e. 0000 UTC equals 7 p.m. EST, 6 p.m. CST, 4 p.m. PST.

UTC	Freq.	Station/Country	Notes
0000	7350	China Radio International	
0000	11595	Democratic Voice of Burma, Armenia	Burmese
0000	6055	Radio Exterior Espana	
0000	4765	Radio Progressso, Cuba	SS
0000	4786	Radio Difusora Roraima, Brazil	PP
0000	3310	Radio Mosoj Chaski, Bolivia	Quehua
0000	4717	Radio Yura, Bolivia	SS
0100	7475	Greek Radio, Greece	GG
0100	7375	Mighty KBC, Netherlands, via Germany	
0100	6185	Radio Educacion, Mexico	SS
0100	7280	Radio Farda, USA, via Germany	Farsi
0100	5025	Radio Rebelde, Cuba	SS
0100	9490	Radio Republica, via France	SS
0100	7335	Radio Romania International	
0100	6180	Radio Nacional Brasilia, Brazil	PP
0100	11765	Super Radio Deus e Amor, Brazil	PP
0100	9780	Voice of America, Sri Lanka Relay	
0100	4955	Radio Cultural Amuata, Peru	SS
0100	4985	Radio Brazil Central	
0200	9420	ERA, Greece	Greek
0200	9550	Islamic Republic of Iran Broadcasting	
0200	11710	Radio Argentina al Exterior	
0200	5910	Radio Japan	JJ
0200	7460	Radio Pyam e-Doost, via Moldova	Farsi
0200	9855	Voice of America	Tibetan
0300	9505	Adventist World Radio, via Austria	Farsi
0300	5980	Channel Africa, South Africa	
0300	7120	Radio Hargeisa, Somalia	Somali
0300	3320	Radio Sonder Grense, South Africa	Afrikaans
0300	4055	Radio Verdad, Guatemala	SS
0300	7205	Radio Omdurman, Sudan	AA
0300	9530	Trans World Radio, Swaziland	Vern
0300	4775	Trans World Radio, Swaziland	Vern
0300	7360	Vatican Radio	
0300	4930	VOA Relay, Botswana Relay	
0300	9515	Voice of Turkey	
0300	7506	WRNO, Louisiana	
0300	5950	Voice of Tigray Revolution, Ethiopia	Tigrinya
0400	6050	HCJB, Ecuador	SS
0400	9470	Deutsche Welle, Germany, via Rwanda	
0400	9625	Radio Fides, Bolivia	SS
0400	4950	Radio Nacional, Angola	PP
0400	6010	La Voz de Concencia, Colombia	SS
0400	4990	Radio Apinte, Suriname	Dutch
0400	4790	Radio Vision, Peru	SS
0400	4960	Voice of America, Sao Tome Relay	
0400	6010	Radio Inconfidencia, Brazil	PP
0500	7275	Radiodifusion Tunisienne, Tunisia	AA
0500	9820	Radio 9 de Julho, Brazil	PP
0500	11675	Radio New Zealand International	
0800	9700	Radio New Zealand International	
0800	1100	Radio Nikkei, Japan	JJ
0800	5930	Radio Rossii, Russia	RR
0900	4810	Radio Logos, Peru	SS
1000	3330	Ondas del Huallagua, Peru	SS
1000	9625	Radio Japan	
1000	5039	Radio Libertad de Junin, Peru	SS
1000	6140	Radio Twantinsuyo, Peru	SS
1000	4700	Radio San Miguel, Bolivia	SS
1000	3905	Radio New Ireland, Papiua New Guinea	Tok Pisin
1100	9870	All India Radio	Hindi
1100	4835	No.Territory SW Service, Australia	
1100	2325	No.Territory SW Service, Australia	
1100	5875	Radio Thailand	Khmer
1100	3925	Radio Nikkei, Japan	JJ
1100	5020	Solomon Islands Broadcasting Service	
1100	4940	Voice of the Strait, China	CC
1100	4750	Radio Repubik Indonesia	II
1100	3260	Radio Madang, Papua New Guinea	Tok Pisin

UTC	Freq.	Station/Country	Notes
1100	4755	The Cross Radio, Micronesia	
1100	5765u	Armed Forces Network, Guam	
1200	15195	Bangladesh Betar	
1200	5915	Myanmar Radio	Burmese
1200	6100	Radio Rossii, Russia	RR
1200	13805	Voice of Russia	Mandarin
1200	3325	Radio Republik Indonesia	II
1200	17660	Radio France International, via S. Africa	FF
1300	9410	China National Radio	CC
1300	9520	China National Radio	CC
1300	11510	Denge Kurdistan, via Moldova	Kurdish
1300	9335	Radio Free Asia, No. Marianas Relay	Burmese
1300	9950	Nippon no Kaze, Taiwan to N. Korea	KK
1300	7405	Radio Marti, USA	SS
1300	9580	Radio Australia	
1300	17520	Radio Pakistan	
1300	11640	Radio Taiwan International	Mandarin
1300	12055	TWR-India, via Russia	Bangla
1300	9526	Voice of Indonesia	
1300	9435	Voice of Korea, North Korea	
1300	11530	Radio Pakistan	Tamil
1300	9965	Radio Australia, via Palau	
1400	12065	Radio Australia	
1400	15330	Radio Veritas Asia, Philippines	Urdu
1400	15180	RFE/RL, via France	Turkman
1400	11990	VOA, via Russia	Mandarin
1400	7490	WWCR, Tennessee	
1400	13845	WWCR, Tennessee	
1499	15575	KBS World Radio, Souh Korea	KK
1500	15506	Bangladesh Batar	Urdu
1500	15660	Channel Africa, South Africa	Swahili
1500	17545	Islamic Republic of Iran Broadcasting	
1500	15320	Radio Veritas Asia, Philippines, via Vatican	Tagalog
1500	17690	Voice of America, via Vatican	
1500	11710	Voice of Korea	
1500	15550	WJHR, Florida	
1599	21540	Radio Kuwait	AA
1600	15150	Adventist World Radio, Austria Relay	Farsi
1600	15205	BSKSA, Saudi Arabia	AA
1600	15345	Radio Cairo, Egypt	
1600	13735	Radio Rossii	RR
1600	15485	Radio Taiwan International, via France	
1600	11765	Sound of Hope, Taiwan	Mandarin
1600	15265	VOA, Botswana Relay	Swahili
1700	12095	BBC, Seychelles Relay	
1700	15255	Channel Africa, South Africa	
1700	15385	ESAT Radio, via Bulgaria	Amharic
1700	15570	Vatican Radio	
1700	15120	Voice of Niigeria	
1800	15400	BBC, Ascension Island Relay	
1800	17795	BBC, England	
1800	15250	Far East Broadcast, Assn. via Ascension	FF
1800	17850	Radio France International	FF
1800	9840	WHRI, Indiana, via South Carolina	
1900	11670	All India Radio	
1900	11875	IBRA Radio, via Germany	Vern
1900	15540	Radio Kuwait	
1900	15140	Radio Sultanate of Oman	AA
1900	11635	Radio Exterior Espana, Spain	
1900	15630	Voice of Greece	GG
1900	15825	WWCR, Tennessee	
2000	11610	Adventist World Radio, via Germany	AA
2000	11730	Adventist World Radio, via France	Yoruba
2000	11820	BSKSA, Saudi Arabia	AA
2000	11750	Islamic Republic of Iran Broadcasting	
2000	11995	Radio France International	FF
2000	15190	Radio Africa, Equatorial Guinea	
2000	11765	Radio Algerienne, via France	FF
2000	11730	Radio Belarus	SS
2000	11850	Radio Japan, via Madagascar	FF
2000	11745	Radio Romania International	
2000	11885	Voice of America, Sao Tome Relay	
2000	9690	Voice of Nigeria	Hausa
2000	11840	Voice of Vietnam, via Germany	VV
2000	11625	Vatican Radio	
2000	11735	ZBC Radio, Zanzibar	Swahili
2100	7550	All India Radio	
2100	11720	BSKSA/Holy Qur'an, Saudi Arabia	AA
2100	7285	China Radio International, Albania Relay	
2100	11940	China Radio International	Bulgarin
2100	11865	Deusche Welle, Rwanda Relay	
2200	15525	HCJB, Australia	Mandarin
2200	12080	Islamic Republic of Iran Broadcasting	AA
2200	6060	Radio Agentina Exterior	SS
2200	11855	Radio Aparecida, Brazil	PP
2200	5860	Radio Farda, USA, via Sri Lanka	Farsi
2200	7450	Radio Makedonias, Greece	Greek
2200	11780	Radio Nacional Amazonia, Brazil	PP
2200	7450	Radio Romania International	
2200	9850	Radio Tirana, Albania	
2200	9465	Voice of Russia	
2300	6160	CKZN, Canada	
2300	5940	Magadan Radio, Russia	RR
2300	8989u	Pescador Preacher, Nicragua	SS
2300	4915	Radio Daqui, Brazil	PP
2300	4915	Radio Difusora Macapa, Brazil	PP
2300	9720	Adventist World Radio	Tibetan
2300	11765	Super Radio Deus e Amor, Brazil	PP
2300	9930	WTWW, Tennessee	
2300	5005	Radio Nacional, Equatorial Guinea	SS
2300	4319u	Armed Forces Network, Diego Garcia	
2300	9645	Radio Bandeirantes, Brazil	PP

Public Safety Monitoring Gets Its Just Due @ Last Month's Consumer Electronics Show in 'Vegas

Exciting New Scanners Expand Uniden's HomePatrol Family

BY GORDON WEST, WPC6NOA/WB6NOA

Last month at the Consumer Electronics Show <http://www.cesweb.org> in Las Vegas, Uniden unveiled two new scanners — its great new BCD436HP and BCD536HP — with a companion app that will advance the consumer's HomePatrol-1 into a mobile and handheld scanner for the *serious* public-safety monitor, **Photo A**.

"Several years ago, *[Popular Communications']* CES coverage included our new [at the time] HomePatrol-1 table-top scanner," Paul Opitz said, **Photo B**. He is *the main man* in Uniden marketing. "Any scanner listener who knows their ZIP code could easily and quickly be listening to the local action!"

*(**WATCH and LISTEN:** To Paul Opitz detail the features of the BCD436HP and BCD536HP scanners in a full-length video at <http://bit.ly/18dAtFt>, **Photo C**. – WPC6NOA)*

***Photo A.** Uniden's new BCD536HP, left, and BCD436HP scanners take the HomePatrol concept to a whole new dimension. New features and integration opportunities are all over this radio! (Internet screen grab)*

You Asked for It

Opitz noted that people *really into* public-safety monitoring wanted a traditional-size scanner product for rack or mobile-unit mounting. Those same scanner users also wanted a true handheld HomePatrol-type scanner, plus the added capability of programming local systems into the radio, thus keeping the programming up-to-date as systems changed in response to narrow banding, re-banding, and system upgrades.

The scanner information database provided by Radio Reference.com <http://www.RadioReference.com> is a great avenue to take to find thousands of frequencies, but the die-hard scanner operator may know about local system "secret channels" and want to program them from the front panel.

Narrow banding and re-banding is in full swing, with APCO Project 25 Phase II taking analog trunking and digital channels to *a new high.*

New Uniden Handheld with 'TrunkTracker V' Technology

The new Uniden BCD436HP and BCD536HP scanners

***Photo B.** The wildly popular HomePatrol-1 tabletop scanner was the catalyst for Uniden's newly-designed and upgraded public-safety scanners. (Internet screen grab)*

**CQ Contributing Editor at Large*
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e-mail: <wb6noa@cq-amateur-radio.com>

and their companion app were on display at CES.

"Our new BCD436HP is the first scanner to incorporate the HomePatrol-1's ease of use in a traditional handheld scanner," Opitz said. To get started with this new scanner, all it takes is entry of your local ZIP code, and the new handheld springs to life.

"The four gigabyte micro SD card holds the entire USA and Canadian database of radio systems, with plenty of room for recording hours of off-air audio," he said. The alert indicator can also flash in a user-set color and pattern when a specific channel becomes active.

Opitz calls this new digital technology reception "TrunkTracker V" to support P-25 Phase I and II, X2-TDMA — TDMA Motorola, EDACS, and LTR trunked radio systems, **Photo D**.

The new Uniden handheld includes a band-scope, trunked system analysis, RF power plot, and EDACS LTR location finder. Simply choose the type of channels you want to hear — police, fire, paramedics, and so on — enter your zip code, and the scanner will do the rest. You can also tie it into a GPS receiver in case you are driving cross-country and always want the scanner to stay tuned to local radio activity.

I especially like the "Close Call®" RF

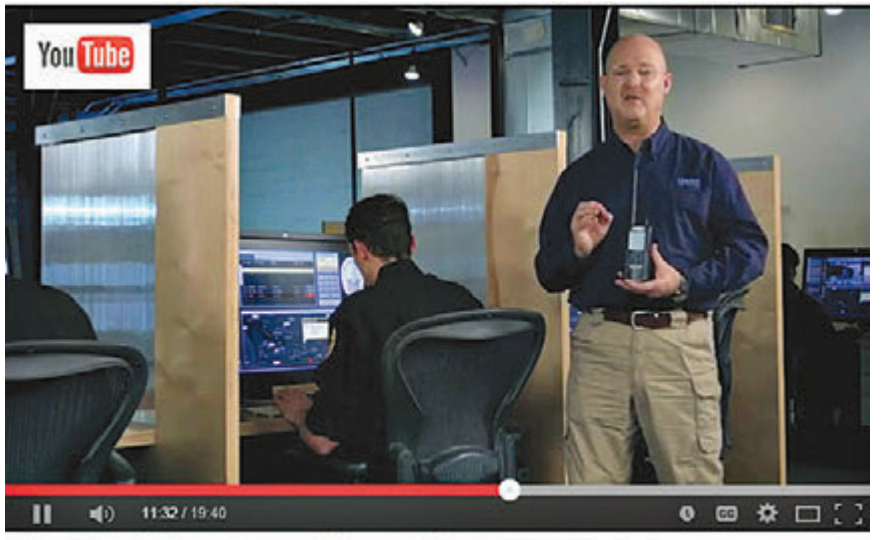

Photo C. *Uniden's Paul Opitz gives a 20-minute seminar on the new '436 and '536 in this online video at <http://bit.ly/18dAtFt>. Details about the new handheld scanner begin at 9:58. The new mobile unit is described in detail beginning at 14:53. (Internet screen grab)*

Photo D. *Uniden expects the retail price of the new BCD436HP to be $499. (Internet screen grab)*

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Photo E. *Programming the mobile Uniden BCD536HP is a snap. Watch and listen as the scanner goes from shipping box to scanning in just minutes at <http://bit.ly/184C4AD>. Uniden expects the price to be $599 (Internet screen grab)*

Photo F. *Read the full Uniden BCD436HP and BCD536HP manuals at <http://bit.ly/18PN4lU>. (Courtesy of Uniden)*

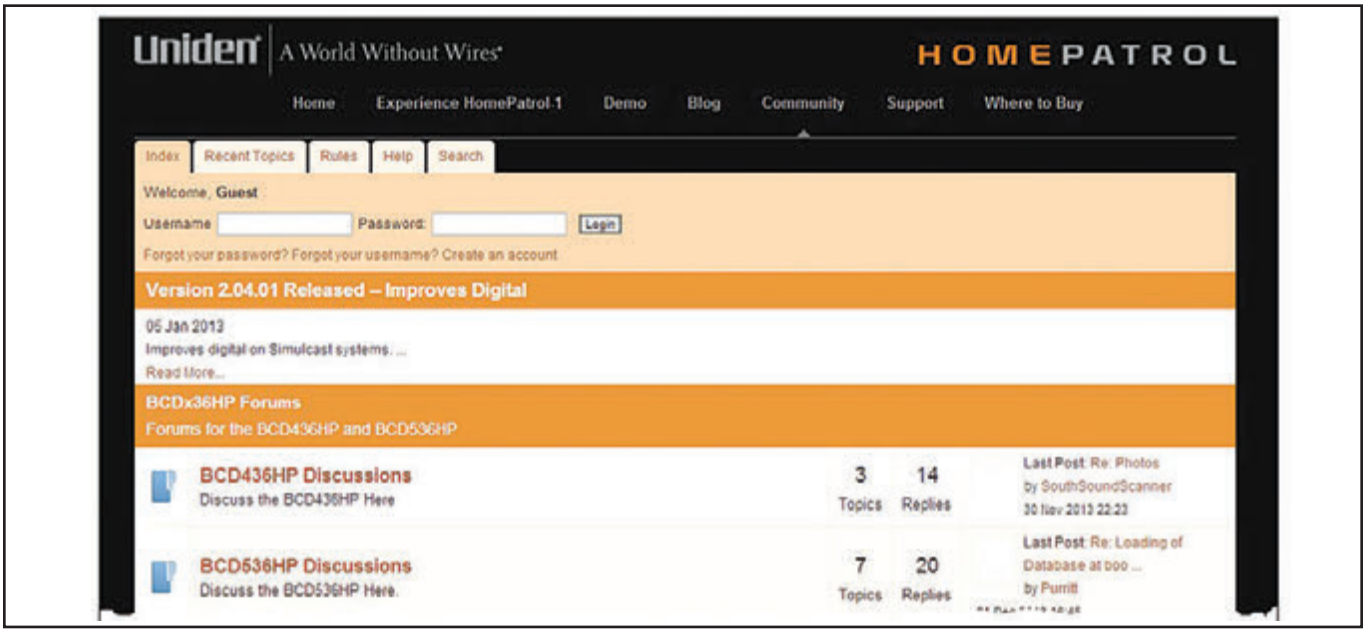

Photo G. *For even more information and insight, Paul Opitz suggests readers visit the Uniden HomePatrol discussion page at <http://bit.ly/1bb1fSa>. (Internet screen grab)*

Photo H. *The Uniden website reflects on the HomePatrol-1 scanner and how the BCD436HP and BCD536 have taken the concept to a whole new level. (Internet screen grab <http://info.Uniden.com/NewScanners>.)*

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The new bhi 10 watt DESKTOP speaker has a 4" bass driver and 1" tweeter - Digital rotary volume and filter level controls - Separate stereo line-in and speaker level inputs - Headphone socket - Audio & LED indication of filter function - Audio overload- Sleep mode - Noise reduction 9 to 35dB - 12V DC to 18V DC power (2.5A peak) - Weight 3.6lb, dims 8"(h) x 6"(d) x 6.3"(w) - Supplied with fused DC power lead, 3.5mm audio plug lead & user manual

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Photo I. *A checklist of features in Uniden's newest handheld and mobile scanner shows the breadth of coverage these great units provide. (Internet screen grab)*

capture, that lets the scanner find nearby transmissions and its ability to program this mode without interrupting on-going scanner activity.

And for those of you in rural areas who still use two-tone paging, this handheld has you covered.

The handheld runs on three AA batteries and Opitz said it should play for up to eight hours on a set. I like the feel of this unit with its rubber sides and rubber bottom impact areas, in case it takes a trip down to the deck.

Mobile Scanner: Integration Opportunities Galore

Of course, the mobile scanner has all the same features as the portable scanner, and both units have *even more* features than the original HomePatrol-1. *(**WATCH and LISTEN:** To the Uniden BCD536HP mobile scanner on the air within minutes of being taken out of the box, **Photo E**, at <http://bit.ly/184C4AD>. – WPC6NOA.)*

Both scanners include Sentinel Software that will make the database and firmware updates simple. This software will allow you to create, edit, and manage your favorite channel list. It is USB-PC cable-ready to enhance managing all of the channels you have in memory. And, like the handheld, you can also program channels from the front panel, too.

Much More Information

Uniden expects the retail price of the handheld BCD436HP to be $499, and the mobile BCD536HP, $599. I talked Paul into giving us some nice sites to learn more about these two new scanners and the software app.

- BCD436HP and BCD536HP Manual, <http://bit.ly/18PN4lU>, **Photo F**
- The Uniden Home Patrol Discussion Page, <http://bit.ly/1bb1fSa>, **Photo G**
- Additional Uniden '436 and '536 information, <http://info.Uniden.com/NewScanners>, **Photo H**

Anxious Anticipation

No, at this writing I don't have one of these new scanners *yet.* I work the original HomePatrol-1 version regularly, so I can hardly wait for some hands-on time with the new mobile '536 and the new handheld '436 Uniden scanner receivers! **Photo I**.

BY RICHARD FISHER, KPC6PC/KI6SN

Getting Started in VHF-UHF Scanning Can Be Free and Easy as 'A, B, C, D!'

In this month's *Internet Radio* we're going to get serious about VHF-UHF scanner listening. There is a fascinating world of police, fire, aviation, governmental, radio amateur, weather, and so much more if you've got *a way to get there.*

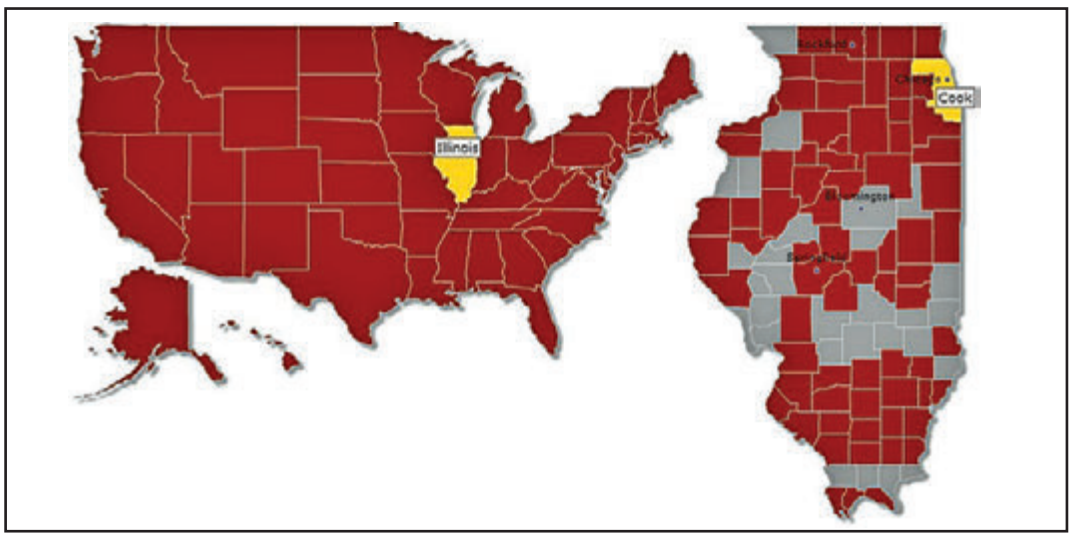
*1940 Wetherly Way, Riverside, CA 92506
e-mail: <CQPlusDigital@gmail.com>

Vagaries in wave propagation generally do not play a major role in "band conditions," so you'll find action in this part of the spectrum 24/7/365.

If you've never ventured into scanning, and would like to listen before making an investment in gear, there are portals on the Internet that will get you hooked up to scanners all over the United States.

Of course, there's no substitute for that radio

STEP A. *Link to the Broadcastify.com website at <http://www.broadcastify.com> and click on "Listen." We've circled it in yellow in this image to make it easy to find.*



STEP B. *Your next screen will look like the image on the left — the map of the United States. For our demonstration, we've chosen the Chicago Police Department as our target scanner. So we click on Illinois. That takes us to the image on the right: a map of Illinois segmented into counties. Chicago is in Cook County, so we click on it.*

STEP C. *Clicking on Cook County takes us to a list of all the scanners online from there. There are a lot! We scroll down to "Chicago Police Department" — which we've circled in red on the left. Click on the small triangle on the far left to activate the audio feed. Notice the number 409 on the right, also circled in red. This is a count of how many people are currently listening to this scanner feed. That's a pretty good indication that we're in for some urban police action.* →

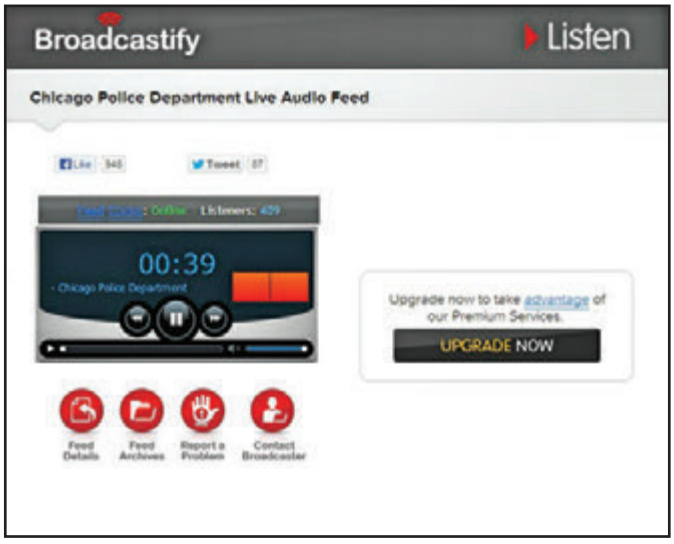
chattering away in the background, or hanging on your belt as you move around town — but "Internet Radio" offers you a no-cost way to test drive the spectrum before buying your gear.

To try things out, all you need is a computer or smartphone with audio and Internet access. Period. For our example, you'll get there by linking to Broadcastify.com <http://www.broadcastify.com>, a website that links users to 3,000+ VHF-UHF scanners across the country. Depending on your device's capability you may need to use another service such as TuneIn, Scanner Radio Deluxe, or other app. Just Google them and you're sure to find one that works.

With Broadcastify, simply follow our four-step process. There are no programs to download, accounts, or passwords to create, or antennas to put up.

STEP D. *Clicking on the triangle noted in STEP C brings you to this screen: Chicago Police Department Live Feed. There is a running clock ticking off the amount of time you are logged in. At the moment of this screen shot, we had been on the CPD site for 39 seconds. The orange blocks to the right of the clock bounce up and down with the modulation of the radio signals from the dispatchers and police officers. The bar below it to the right is the volume control. We have it cranked up to full volume and adjust the overall volume via our laptop.* →

Imagine: Listen to the fire department in your town, or the town *you used to live in.* Or, you've just heard on the news there is a major incident going down that's got national attention. The Boston Marathon terrorist bombing, for example. Sure, you can get a TV or radio reporter's perspective. But how about tuning in a Broadcastify.com police or emergency service scanner and hear the action first hand? *It's really as easy as A, B, C, D!*



New Members: *Monitoring Station* Program

Here are the newest station monitors granted a station identification sign, authorized a Certificate of Registration, and welcomed to the *Pop'Comm Monitoring Station* program.

KPC and DX Prefixes

They are listed by name, station identification sign, and monitoring station location.

Donald Osmond, Sr., **KPC4SKT**, Tega Cay, SC; Stephen Piwonski, **VKPC3AIR**, Melbourne, Victoria, Australia; Mark Prock, **KPC5KQQ**, Allen, TX; Joseph Havens, **KPC9ZDA**, Yoder, IN; James Andrews, **KPC8SSS**, Kettering, OH.

WPC Prefixes

Also: Darwin McDonald, **WPC8DMD**, Madison Heights, MI; John Green, **WPC3BV**, York, PA; James O'Neil, **WPC7JSO**, Newsmanstown, PA; Paul Rehkopf, **WPC-8PRR**, Montpelier, OH; Christopher Gearhart, **WPC1HWQ**, South Windsor, CT; Gary Dobbs, **WPC8GD**, Sarver, PA; Michael Fontaine, **WPC4BBS**, Wesley Chapel, FL; Jeffrey Marker, **WPCØFYV**, Raytown, MO; Christopher Jerry, **WPC4CJ**, Jensen Beach, FL; William Abshear, **WPC1QDR**, Waynesville, OH; William Harrison, **WPC4SC**, Elgin, SC; John Bishop, **WPC4LPF**, Melrose, FL.

For complete information on the *Pop'Comm Monitoring Station* Program and to join, visit *Pop'Comm Monitors On the Web: <http://popcommmonitors.blogspot.com/>.*

– Jason Feldman, WPC2COD
Director, PCMS Registration
<PopCommMonitor@gmail.com>

BY STEVEN HANDLER, WPC9JXK

Commander Bunny, the Pirate Who Would Be President

With No Apologies, He Takes On the Government and the Human Race

This month we have an interview with long-time shortwave radio pirate Commander Bunny of WBNY fame. Not only will we look at his long pirate radio history, but we get a very interesting glimpse of the "inner Bunny." *(**LISTEN:** To a 6+ minute audio clip from Commander Bunny in a YouTube posting from February 2011 at <http://bit.ly/18jvo41>, **Photo A**. – WPC6JXK)*

***[Handler]** First, the obvious question: What motivated you to become a shortwave pirate broadcaster?*

[Commander Bunny] I have always had a shortwave around, listening to my dad's in the 1950s. During the '60s I heard of *pirate* broadcasters via *Popular Electronics* magazine. That set my imagination soaring and I searched often, and once in a while would find an errant ham playing music. But I felt there was a niche I could fill.

I picked up a Johnson Viking Ranger in 1965 from the wife of a *silent key* [deceased amateur radio operator] neighbor, **Photo B**, and decided after a year of being cautious and waiting that I would have some fun.

I had spent many hours in that neighbor's basement, being *Elmered* [mentored] by him on all things electronic. Plus my dad had been a telegrapher on ships in the 1920s and '30s. So I grew up around it.

e-mail: <Stevenhandler-popcomm@yahoo.com>

***Photo A.** A YouTube posting from February 2011 captures about six-and-a-half minutes of a Commander Bunny broadcast on shortwave. (**LISTEN:** At <http://bit.ly/18jvo41>.) (Internet screen grab)*

Fun has always been what I wanted to have and still to this day remains my motivation. That's fun *for me,* and hopefully fun for people listening. If there is any bigger rush than firing up a transmitter and throwing a signal into the ether, not knowing where it will go, or if you might get caught doing it, then I don't want to know about it.

Then you add to that a program that you created and there isn't a drug on the face of the planet that can give you a bigger adrenaline rush.

*(**NOTE: Photographs C** through **K** were provided to Pirate Radio courtesy of Commander Bunny. – WPC9JXK)*

***[Handler]** I understand that another pirate operator originally used the callsign of WBNY. When did you begin operating WBNY and how were you affiliated with the original WBNY pirate station operator?*

[Commander Bunny] Your understanding is wrong — just another nasty rumor floated by the shadowy figures who have nothing better to do with their time on supposed pirate blogs than throw out lies, gossip, and rumors.

This *rumor* didn't start up until the mid-2000s when I was on the air just about every weekend. This was also about the time that [Internet] *chatrooms* came along and the rumors about every pirate operator or suspected pirate operator became the norm.

If I went around denying or trying to correct all the rumors and gossip about me, I wouldn't have time to broadcast. This is the one part of [pirate radio] that detracts from the fun. Sadly this comes with pirate radio fame. The more famous and well known you become, the bigger target [you are] for the jealousy and snipes.

I've had one especially nasty stalker who has blamed me for everything from FCC busts to deleting one of his posts on a pirate website. *Trust me about this:* Only one person knows who I really am, and that's my drop operator [the person who receives mail and forwards it to the pirate broadcaster].

There are lots of rumors about who they *think* I am; who they *think* I sound like and some claim they know — but none of these [guesses have been] even close.

The first rule of pirate radio is *trust no one*. And I haven't.

WBNY has been on the air in one form or another every Easter since 1966. Were I any of the people the *know-it-alls* claim, I'd have been shut down long ago.

While the Internet has brought reception reports almost immediately, it has also given a big voice to the gossips and rumor mongers. Believe nothing you read on blogs and websites. These are just

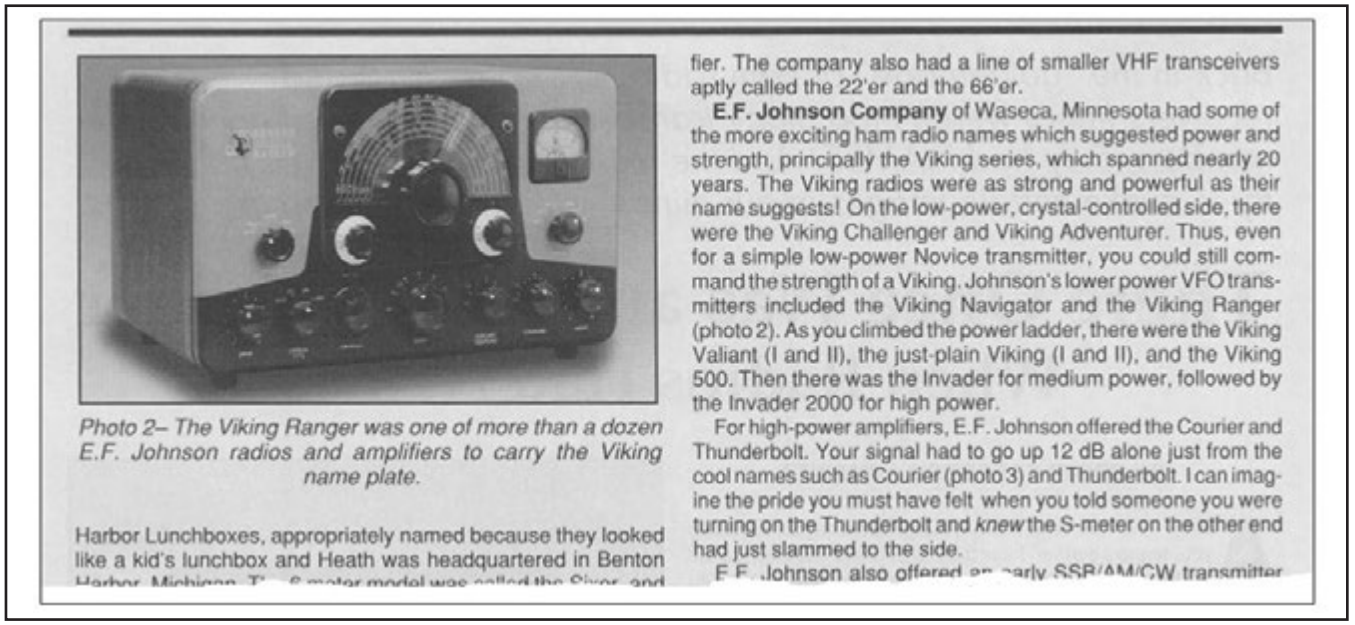


Photo 2– The Viking Ranger was one of more than a dozen E.F. Johnson radios and amplifiers to carry the Viking name plate.

Harbor Lunchboxes, appropriately named because they looked like a kid's lunchbox and Heath was headquartered in Benton

fier. The company also had a line of smaller VHF transceivers aptly called the 22'er and the 66'er.

E.F. Johnson Company of Waseca, Minnesota had some of the more exciting ham radio names which suggested power and strength, principally the Viking series, which spanned nearly 20 years. The Viking radios were as strong and powerful as their name suggests! On the low-power, crystal-controlled side, there were the Viking Challenger and Viking Adventurer. Thus, even for a simple low-power Novice transmitter, you could still command the strength of a Viking. Johnson's lower power VFO transmitters included the Viking Navigator and the Viking Ranger (photo 2). As you climbed the power ladder, there were the Viking Valiant (I and II), the just-plain Viking (I and II), and the Viking 500. Then there was the Invader for medium power, followed by the Invader 2000 for high power.

For high-power amplifiers, E.F. Johnson offered the Courier and Thunderbolt. Your signal had to go up 12 dB alone just from the cool names such as Courier (photo 3) and Thunderbolt. I can imagine the pride you must have felt when you told someone you were turning on the Thunderbolt and *knew* the S-meter on the other end had just slammed to the side.

E.F. Johnson also offered an early SSB/AM/CW transmitter

Photo B. *A Johnson-Viking Ranger transmitter, given to him by the widow of an amateur radio operator neighbor, opened the transmitting world of pirate radio to Commander Bunny. (With permission, courtesy of CQ magazine)*

Photo C. *Here's an example of a WBNY Relay Service QSL. (Unless otherwise noted, images courtesy of WBNY and Commander Bunny)*

people who want attention because they can't broadcast or if they do anything worth listening to.

[Handler] *When you began operating WBNY what were you trying to accomplish, and along the way have your goals changed?*

[Commander Bunny] The first goal was not to get caught. Remember this was the '60s and the FCC agents of that day were the hard-core, no-nonsense guys you didn't want banging on your door at 2 a.m.

Seriously, I just wanted to have some fun, poke fun at the government, poke fun at the way human beings — [all of whom the Commander considers] *monkey-boys, ape-humans* — act and treat each other.

For the first 10 years of doing WBNY, at least the Easter broadcasts, I had no way of knowing if anyone even heard, except after I'd go off the air, someone would answer [on the airwaves] and say *that was funny* or *thank you!* It was enough to keep me trying.

Remember back in the day, you'd have to find a shortwave broadcast station that wasn't on the air and use that frequency before or after they came on the air.

Although 6 to 7 MHz has always been popular, I've gone to 4, 9, 13, and 20 MHz and even below and above the Citizens Band.

It was during the '70s that newsletters started popping up with some serious pirate listeners posting logs. But these newsletters were only published every other month, or so. But it at least gave you an idea that more than a couple people were hearing you.

The '80s brought not only monthly newsletters — *Free Air* and *The Association of Clandestine Radio Enthusiasts* (ACE) — but dial-up BBSs.

The '90s saw the Internet start to make some inroads — *Free Radio Network* — and establish some real two-way communications between listeners and operators.

The 2000s started off pretty good, but by 2010 it had attracted a lot of negative people who wanted attention at any cost. Pirate radio is about creating programs and broadcasting them, not blogging and posting.

It does appear that a lot of the negative folks have found other shiny objects to keep them occupied now ... but the mess they left will take a long time to clean up. Your question about *the original* is but one of the false rumors floating about.

[Handler] *You created the character of Commander Bunny, a rabbit that heads the "Rodent Revolution." The goal of this revolution apparently is having the rodents seize control of the planet from the human population — which you often refer to as "ape-humans," "monkey boys," and similar terms. What was your inspiration for the character of Commander Bunny?*

[Commander Bunny] I'm not sure I had an inspiration. Darwin? Making fun of the evolution deniers who can't see that all humans act like monkeys, most of the time?

Can anyone deny that things wouldn't be better if the rabbits ran things? I took a lot of what was in *Planet of the Apes* in the late '60s and incorporated that story line into the *Commander Bunny* persona.

Instead of monkeys running things it would be rabbits (rodents) and well, yeah, humans would be the slaves. Face it, humans — as we know them today — have only been

Photo D. *Commander Bunny is a candidate for U.S. President — at every opportunity.*

around for maybe 1 million years. Our planet is 4.5 billion years old. Who was running the show before humans? Who will run it when humans are gone? And there is no question humans will be gone at some point.

Has there ever been a more arrogant creature on Earth than people? Next logical *(?)* step would be for the rabbits — *rodents* — to take over.

[Handler] *At some point you came up with the idea of your character, Commander Bunny, running for U.S. President. When did you begin this story line and what was your inspiration for this idea?*

[Commander Bunny] It was 2004 and there were 12 — *count 'em* — 12 candidates on the election ballot. If John Kerry and Ralph Nader could run for President why not *Commander Bunny*? I had bumper stickers and QSLs made up for the event and the feedback was awesome. I don't know how many votes I got. I'm pretty sure the machines were rigged. But the shortwave community thought it was great! I got a lot of fun, positive feedback. And I still do to this day.

[Handler] *What were you trying to accomplish with the Commander Bunny for President story line and have you accomplished those goals?*

[Commander Bunny] My goal was to establish that *anyone* can run for President. I have run for President ever since. Hey, going in the direction we're going, my candidacy isn't looking so bad now, is it?

It was also a reason to have another type of QSL printed up. I really believe that a fun, different QSL makes the pirate radio experience worthwhile.

I know the newer generation of pirates do e-QSLs, but there isn't anything better than sending off to some strange post office box with your report, then waiting for a QSL to show up — a [card] you can hold in your hand, show your friends, and hang on your listening shack's wall. I fear the *real-QSL* event is being lost to technology.

[Handler] *Your broadcasts tend to have very high production values. Once you have envisioned the concept or theme of a show, please describe the artistic and production process it takes to move from an idea or concept to a final broadcast.*

[Commander Bunny] Sometimes the inspiration comes quickly, the elements of the show fall into place and I can put together a 45-60 minute show in two to four days [consisting] of six-hour sessions at the mixer and computer.

It may sound flawless and effortless, but *trust me*: It is a labor of love. I have a Yamaha 8-in, 4-out stereo mixer and lots of years' worth of parody songs and silly commercials. I work out the basic idea of the show in my head. Then, as I'm putting it together, all the pieces seem to fall into place. *Monkey* songs just seem to show up from listeners who send them or I'll find it at an old record store or at a yard sale.

[Handler] *Once you have your script written, about how long does it take for you to create one of your programs?*

[Commander Bunny] I usually get the idea in my head, think about what I'm going to do, write a basic script, then sit down at the mixer and keyboard. I'll get the thoughts about the program written down — script and technical notes — and the rest of it usually ***(NOTE:*** *usually)* comes after that. Sometimes it's like pulling teeth, but that's the exception, not the rule.

[Handler] *Are your program recordings done in a single take, or do you use multiple takes, and if so, about how many on average for a show?*

[Commander Bunny] The voice part of it is done in a single take. I want a flow to the program. Doing edits interrupts [that flow]. Remember, I've

Photo E. *This WBNY Relay Service QSL celebrates the recent FCC shutdown.*

Photo F. *A commemorative Commander Bunny for President coffee mug hails the Rodent Revolution and declares "America Needs a Hare Transplant."*

Photo G. *The WBNY Relay Service relays "the best of pirate for MWNT (monkeys with no transmitters)."*

Photo I. *Here's one of the official QSLs of the WBNY Relay Service.*

Photo H. *WBNY is the broadcast conduit for CHKN — Chickenman, FOWLEST Shortwave Pirate Ever!*

Photo J. *WCS, The Collective Soul Station, is on the air thanks to Commander Bunny and WBNY.*

already written the script in my head, then written it down so the script is more like notes. It's the editing that takes an enormous amount of time and sweat.

[Handler] *Your broadcasts have made use of satire, one of the oldest and most sophisticated forms of humor. Do you write all of your material or do you have someone who assists you?*

[Commander Bunny] Nope, there's no one but me to blame. I read newspapers, watch the news, follow politics, see posts on the FRN (Free Radio Network) and see how silly people act. Satire is simply observation of the obvious. I suppose it's a point-of-view issue. You see the United States Congress in session (when it isn't on vacation, in recesses, or during a shutdown) and I see monkeys acting (OK, so it isn't an "act") like idiots. Do you watch *CSPAN* or John Stewart on *The Daily Show?* **Photos L** and **M**.

[Handler] *In the use of satire, how do you balance your right to free speech against airing material that some in your audience may find offensive?*

[Commander Bunny] The only thing that any of my listeners have found offensive is my use of the *monkey-boy* and *ape-human* phrases. I am still amazed by how much those terms made people angry.

So someone who calls himself *Commander Bunny* calls you a *monkey-boy* and you're going to take it seriously? *Really?* That's too funny!

Anyone offended by language should probably find a nice PBS station, not pirate shortwave radio. There's not a lot of vulgarity in my shows, but you probably won't be hearing any of the WBNY programs on the local *Baptist Hour.*

[Handler] *I understand that beside WBNY and WBNY Relay Service, you have operated many other pirate radio stations with varying formats and persona. Commander Bunny's revolution is now in its second decade. Do you have a desire to try something different and if so, what?*

[Commander Bunny] *Commander Bunny* in his current state is in his *third* decade. WBNY is in its *sixth* decade. The majority of my broadcasts are relays for other pirates now, and I'll continue to do that. I never know what will happen until the urge strikes me.

[Handler] *Throughout your pirate broadcasting career, is there a single broadcast that stands out from the rest as your "best broadcast" and if so, please tell us about that program.*

[Commander Bunny] That's kind of like asking a parent — with their kids standing there: *Which child do love the most?* They have all been fun. Fun to think up, fun to write, fun to produce, and fun to air.

It has been fun to send QSLs and fun to hear the reaction of the QSL packages. I usually send a super QSL package to about the fourth confirmation, but if one hits me funny, I'll send them the whole tote-bag (spoof on PBS), coffee mug,

Photo K. *This QSL is for WHOF — Hall of Fame Radio — which broadcast through the facilities of the WBNY Relay Service.*

pens, magnets, *yadayadayada* as my appreciation of their getting the joke!

OK just thought of my favorite show: It was the *Good Monkey* show. I had asked both on the air and through the FRN (Free Radio Network) for people to send me audio with them saying, *I'm a Good Monkey.* I got more than I ever dreamed of. It took me weeks of daily editing to get them all in, but I did.

I counted 74 different clips of people saying *I'm a Good Monkey* on that show. It showed not only how many people listen but are willing to take the time to cut an audio piece for the show.

My favorite Easter show (that was what got me started with WBNY) is the Easter 2007 show where I incorporated a South Park program about why Easter eggs are painted and why the Pope has a big hat (to hide his bunny ears).

[Handler] *As a pirate shortwave radio operator, have you accomplished all of your goals and if not, what's left?*

[Commander Bunny] Well, since all I ever wanted to do was have fun, I think I have. Each station I have created over the years was for a specific purpose that, at the time, met the goal I wanted.

The one goal that I've accomplished and enjoyed is being a Relay for pirates who don't have transmitters. I've relayed hundreds of pirates who either didn't have a transmitter or just wanted to throw the FCC off by having them pop up from a different part of the country.

I've made many friendships because of the Relay function. A lot of very talented people are out there who have great, creative ideas but have no way of getting them on the air. It has been a ball to provide them with a way to get their programs broadcast. But the underlying goal of having fun has been met and continues to this day.

[Handler] *The licensed shortwave broadcasting community is contracting worldwide. Having been both a participator in, as well as an observer of shortwave radio broadcasting, what impact will the demise of many licensed shortwave broadcasters have on pirate radio broadcasting in North America?*

[Commander Bunny] It's not just shortwave. It's the entire broadcasting world. Have you listened to AM and FM radio lately? Broadcast listenership across the board is declining. Shortwave is a niche group of hardcore, "I love radio*"* kind of person. And the *pirate shortwave radio* kind of person is an even smaller niche, but in my opinion it will continue in spite of what commercial radio does.

The fact that they have a shortwave and listen to it speaks volumes to their commitment to a very unusual hobby. The demise of commercial shortwave should open up some frequencies for pirates.

The typical commercial shortwave broadcaster splatters for 50kHz +/-, so pirate listeners aren't going to miss them in the least. It is funny how the FCC will chase 50-watts-or-less pirates but can't seem to find the time to go after the 100-kilowatt "*splatter-the-entire-band"* commercial stations.

As long as they make shortwave receivers, there will be shortwave listeners and then a niche of folks who find pirate shortwave even more appealing than listening to commercial preachers begging for money or hams arguing over who *had the frequency first.*

During the era around 2005, there were some people who complained about the low-power pirates and *who wants to listen to a low-power pirate?* These are the Generation X, *gimme everything right now* group, that has since moved on to something less challenging than pirate radio.

The romance of hearing a low-power pirate broadcaster is lost on that generation. It wants FM high-powered, *don't make me work for stations.*

Pirate radio is not for them. But to the point, the demise of the big 100-kilowatt shortwave stations will make low-power pirate reception more attainable.

Photo L. *Commander Bunny tunes into John Stewart and The Daily Show for perspective on the content of WBNY broadcasts. (Internet screen grab <http://www.thedailyshow.com>)*

[Handler] *What role in pirate radio broadcasting do you envision for yourself in the future?*

[Commander Bunny] I'd say: staying at least one step ahead of the FCC!

I don't see me doing anything different in the future than what I've done in the past. It has worked for six decades. *Why stop now?*

I don't try to lead anyone or make anyone believe anything they otherwise wouldn't believe. My role isn't to be a leader in pirate broadcasting. My role is to practice my craft and have fun doing it.

WBNY, Commander Bunny, Radio Free Speech, and all the other stations were done as entertainment and for

Photo M. *The machinations of Congress, as covered by CSPAN, is a source of inspiration for Commander Bunny.*
(Internet screen grab <http://www.c-span.com>)

fun. Whatever role that has in pirate radio, then that's what I'll be doing. *Nothing more, nothing less.*

The people trying to bend public opinion or attempting to mold a movement or tell people they know who a pirate is, or have some tidbit about an operator — they have another agenda. It's an agenda I don't share.

These are the same people who sit next to you in a magic show and after the trick is performed have to tell you *how it's done.* I don't want to know how it's done, I want to be tricked into thinking it's something magical.

Pirate radio, at its very essence, is magic. It's some crazy guy (or gal) pretending to be some made-up personality, convincing you of it, putting that pretense on a shortwave transmitter, making it travel through the air and into your radio. *If that isn't magic, then nothing is.*

All of these people who have Internet blogs who pass along the latest rumor, or spread gossip, are in it for themselves *and no one else.*

My respect goes out to all the operators who create shows, then broadcast them for the entertainment of all. It's a shame someone would attach themselves to something that is so much fun, and have nothing but negative comments about who does it and how they do it.

Why would you set up a so-called pirate blog to *out* pirates' identities? How does that make you a part of pirate radio?

Pirate radio has been a wonderful outlet for my views and comments on society and government. It was a natural outlet for my technical abilities.

I am not a ham operator and never have been. I'm not going to agree to let the government come into my home, anytime they want, to *inspect* anything. But I know code, transmitters, and antennas.

Regulating radio waves is like telling people how much sunlight they can use. If you aren't interfering with anyone and operating a low-power HF station on a frequency that is not in use, I don't see what the problem is.

Like the guy on a deserted stretch of Interstate: No is one around. See how fast your car will go. *Who does that hurt?* This *Mama-government* mentality has led to what?

- Well, the latest government shutdown
- AM-FM and shortwave broadcasting collapsing

[It's just that] the average person can't tell the difference between the shutdown and when [the government isn't] working.

I don't *hate* government. I just think there is *way too much* of it. Tea Partiers are left-wing radicals (yes, I said *left wing*), in my view.

In Appreciation

I wish to thank Commander Bunny for taking the time and effort needed for this interview. – WPC9JXK.

'Commander Bunny' Pirate Stations

Alphabetically, here is partial list of pirate radio stations operated or relayed by Commander Bunny.

- Al Fansome Eulogy Radio – Ode to Kevin Miller
- All Average Music Radio
- Cathedral of Chris Radio
- Cell Phone Radio
- Comet Project Numbers Station Spoof
  (LISTEN: *<http://bit.ly/hpaVyv>)*
- Dobson Monkey Radio
- Elvis Annual Birthday Radio
- Family Radio spoof station
- Hitch Hikers Guide to the Galaxy Radio
- KDRM – K drum radio (all drums all the time)
- KRAP Radio – all crap all the time
- Marine Corps Birthday Radio
- Mouth of Mohammed
- My New Underpants
- Radio 9
- Radio Free Speech
- Radio Fudd
- Radio Slack – Church of the Sub-Genius
- Stairway to Heaven Radio
- VOM Voice of Mel (Blanc)
- VOTE – Voice of the Extremist
- We Monkey Radio
- WBNY – Radio Bunny
- WBNY – Relay Service
- WHOF – Hall of Fame Radio
- WJAY – Jay Smilkstein Radio
- WJVR – John Valby Radio
- WMDR – World Monkey Domination Radio
- WRAY – Link Wray tribute station
- WPDR – President's Day Radio
- WRRR – World Rabbit Rapture Radio
- WTKY – Turkey Radio
- WTPR – PT-1 & PT-2 Tire Pressure Radio
- WWWW – All W's all the time
- XENU – Over lord of the Universe (Scientology spoof)

Pirate Radio Listings:
For the latest Pirate Radio loggings, visit <http://www.CQPlusPirates.blogspot.com>.

trail-friendly radio

BY RICHARD FISHER, KI6SN

Measuring Your Location's Antenna Possibilities Without Leaving Your Computer

Space: The final frontier. OK, we're not aboard the *Starship Enterprise.* And for radio amateurs, the frontier of *space* is far from final. Indeed it's ongoing, especially where antennas are concerned. So let's strike that line of thinking.

Many hams who work the high-frequency (HF) bands — 1.8 through 28 MHz — find themselves refereeing a wrestling match with *Lot Space vs. Antenna Length* on the top of the card.

It can be hard to calculate the HLD (height, length, direction) possibilities when you're standing in the middle of your backyard craning to see the tops of a couple of trees with the Sun in your eyes.

"If my antenna goes from *there* to *there* and makes a turn *there,* can I fit an 80-meter dipole on this suburban lot?" is the question. "And what will my spouse think?"

Everyone: To the Web!

We'll leave the last question for another installment of *Trail-Friendly Radio.* As for the antenna and space calculation, though, there's a really cool site on the Internet that can take out almost all the guesswork. It's *FindLotSize.com* and can be used to measure point-to-point distances on virtually any piece of land in the world.

By the way, this program is perfectly suited for the shortwave-antenna builder, as well. *So, SWLer's read on!*

*1940 Wetherly Way, Riverside, CA 92506
e-mail: <CQPlusDigital@gmail.com>

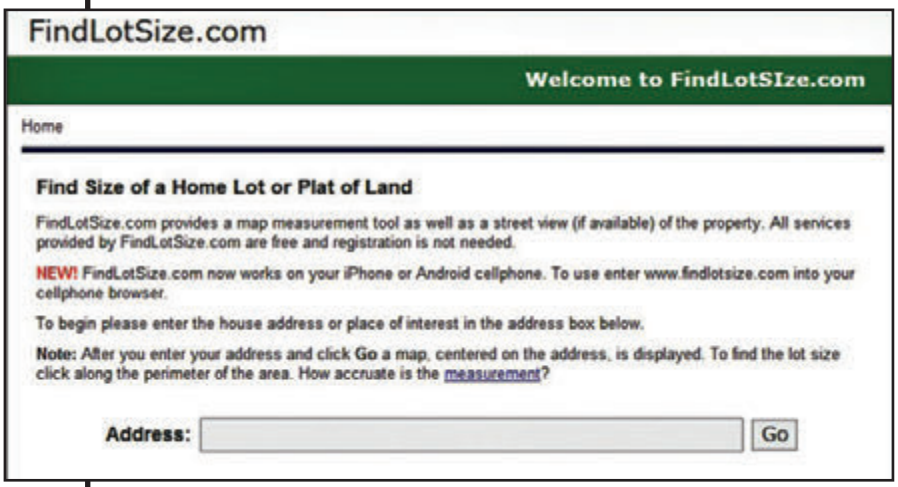


Photo A. *The welcome page at FindLotSize.com features an address box that will "find" the location of the antenna you're thinking about putting up. (Internet screen grab <http://www.FindLotSize.com>)*

Link to <http://www.findlotsize.com> and let the adventure begin.

Upon landing there you'll be greeted with a welcome page that invites you to punch in an address. In your case, the one at which you want to string your antenna, **Photo A**.

Trail-Friendly Calculations

For demonstration, Martha McLean Anza Narrows Park, at 5759 Jurupa Avenue in Riverside, California was typed in. Hit "GO" and you're linked to an aerial view.

This beautiful park in Southern California is a great place for trail-friendly radio operation. The sprawling open space is dotted by towering trees — many providing shade over picnic tables.

This location had been used here several times for portable operation on the higher HF bands — 20 and 10 meters. But what would be the possibilities for working, say, 80 meters from the park now that the summer static has given way to winter quiet?

Knowing the landscape, two giant trees on a gently sloping hill might be perfectly positioned for a dipole cut for 3.6 MHz. Would a half-wave dipole antenna fit? And where would its feed point fall?

By pointing and clicking on the top of each tree, FindLotSize.com — toggled for "Distance" — instantly gave the information needed, **Photo B**. The distance between the map pins is 177 feet — more than enough for a 130-foot-long, 80-meter, half-wave dipole cut for 3.6 MHz. In fact, it could hang in open space with the dipole's end ropes stretching about 24 feet from each treetop. The coax feed line would drop midway — about 88 feet from the base of each tree. All of this calculated from two simple clicks in the FindLotSize.com program. *How great is that?*

For City Slickers

But say we are on a city lot that could in no way accommodate a straight, horizontal low-band antenna. *Time to get creative.*

Let's imagine we're at the address of one of the childhood homes of radio pioneer Edwin Howard Armstong — on the Upper West Side of New York City. The SSB leg of the *CQ World-Wide 160-Meter Contest* is this month — February 21-23 — and we want in on it. But from New York City? *How?*

Turning again to FindLotSize.com, we punched in the legendary EHA's address and consider the possibilities.

Photo B. *By clicking on the treetops shown in this aerial map at Anza Narrows Park in Southern California, it's easy to calculate the distance between them — 177 feet — which is displayed at the lower left of the FindLotSize.com page, provided the "Distance" button above it is clicked. (Internet screen grab)*

Photo C. *Got an antenna configuration that isn't in a straight line? FindLotSize.com handles such distance calculations easily. Put in as many map-pin turning points as you'd like. (See text.) (Internet screen grab)*

Yes, we have trees.

How long is a half-wave dipole for Top Band? Using [468/Freq. in MHz] we see it's 260 feet, or 130 feet per side [468/1.8 = 260; 260/2 = 130].

Hummmm. Starting from the tree in the backyard, **Photo C**, we plot a line to one in the front. Then turn left to a tree across the driveway entry. FindLotSize.com calculates 309 feet — circled in red in our graphic for reference. *Great!*

And imagine the center-fed coax would have had a rather short run to the top floor EHA's bedroom window facing the street. *Perfect!* Who says you can't work Top Band from the inner city?

*(**CAVEAT:** Remember, this is the "site" of the Armstrong house. Obviously, a lot has changed since he lived there as a child. We're only "imagining" the trees were there when he was. And who knows what the house and surrounding area looked like. – KI6SN)*

Add as many map pins as you'd like. The program is perfect for measuring the possibilities of dipoles, long wires, loops, rhombics — in the country or the city. Measure the circumference of your suburban lot. Let your imagination run wild.

But How Do I Calculate Tree Height?

Length is one challenge. Height, another.

Here are four easy steps to calculate the height of any distant object. In our case, a palm tree supporting a sloping G5RV at KI6SN certainly qualifies, **Photo D**.

- Find a stick that is equal in length to the distance from your cheekbone to your fingertips when your arm is fully extended in front of your face. Breaking a branch to proper length will work just fine. *This isn't rocket science.*
- Hold the stick vertical by the tips of your thumb and index finger and put it in front of you with your arm fully extended and parallel to the ground.
- Walk toward or away from the tree until the top of the stick is visually lined up with the top of the tree and the bottom of the stick is lined up with the bottom of the tree. Visually, think of the stick *eclipsing* the tree — like the Moon covers the Sun in a solar eclipse.
- *Viola.* The distance from where you are standing to the base of the tree is equal to the height of the tree. Use a measuring tape if there's one handy. If not, measure the approximate distance between your steps as you walk naturally. Count the number of steps you take to the base of the tree and multiply it by the number of feet and inches each of your steps cover.

Photo D. *A stick cut to the distance between your cheekbone and fingertips when your arm is extended is all you'll need to calculate the height of a tree you're eyeing for a new antenna. It's part of a simple four-step calculation that* works every time. *(See text.) (Courtesy of KI6SN)*

That's all there is to it. ***(WHAT'S UP?*** *The G5RV is 35 feet high in the palm tree at KI6SN. The stick used as the "eclipser" is piece of plastic cord channel 2-feet, 5 1/2-inches long — the distance between my cheekbone and extended fingertips. – KI6SN.)*

Next Steps ...

So, there you have it. If the weather's too crummy right now to do any meaningful outdoor antenna work, link to FindLotSize.com to begin preparing for spring and summer. It's never too early to determine if antenna dreams for your yard or 2014 ARRL Field Day site are going to "measure up!"

BY BRUCE A. CONTI, WPC1CAT

Listening Waters: A Prince Edward Island DXpedition

As in Real Estate, for AM BCB DXers It's 'Location, Location, Location'

Finding a DX location on Prince Edward Island during the November 2013 off-season was a challenge. Most rental properties, especially beachfront vacation cottages of primary interest for saltwater enhanced DX, were closed for the season by the end of October. Fortunately, thanks to the Internet and Google Earth, chances of finding a good spot improved considerably without ever leaving home.

After a successful 2012 DXpedition in Howe Bay on the southeast shore of Prince Edward Island, it was decided to do it again. This prime location provided saltwater paths to Europe, the Middle East, and Africa, along with the Caribbean and South America. Unfortunately the Howe Bay cottage was not available this time. So the search for a new site was underway, beginning with identification of desirable towns on the southeast shore of the island, followed by a Web search for cottage rentals in those towns. Then it was just a matter of checking the listings for properties that were open year 'round.

e-mail: c/o <CQPlusDigital@gmail.com>

PEI DXers. *Members of the 2013 Prince Edward Island AM BCB Listening DXpedition team, from left, were Bruce Conti, Walt Salmaniw, Niel Wolfish, and Brent Taylor — ready for action. (PEI images courtesy of WPC1CAT)*

Once the search was narrowed to just a few still available in November, the sites were located on Google Earth for evaluation. Locations in a compact residential neighborhood, not actually on the beach, or in some cases not even close to being as advertised, were eliminated by Google Earth aerial visits.

What a great tool — something anyone can use to check out locations before ever setting foot on a site or contacting the landlord.

Culling the Choices

The search was narrowed to five potential properties. Now it was time for DXpedition leader and island resident Brent Taylor, VY2HF, to scout the sites with a portable radio to check for noise problems before making any final decisions.

Two properties were selected, but the first contact with a proprietor was negative, as it was not available off-season. Our next best site was on 17 acres of beach front; the Glistening Waters Victorian Farmhouse in Murray Harbour North.

Seventeen acres on the water was a rare find, and if available in November would be even rarer. *Success!* Although the owner had never before remained open so late in the season, she agreed to accommodate us for a week in November. In fact, she was intrigued by our upcoming adventure. *The DXpedition was on.*

It didn't take long for the word to get out about another Prince Edward Island Listening DXpedition either. The farmhouse — from here on dubbed *Listening Waters* — could accommodate four DXers comfortably: The core DXpedition team of Brent Taylor from Stratford, Prince Edward Island; Niel Wolfish, from Toronto; and me from New Hampshire, plus one more. In only a matter of days, Walt Salmaniw, of Victoria, British Columbia, secured the fourth seat. For Walt, an accomplished West Coast North America DXer, this would be his first East Coast DXpedition.

Pre-DXpedition Planning

"Planning ahead is essential anytime multiple operators are involved and when factors of significant travel distance, lodging expense, etc. are in play," recommended medium-wave broadcast band DXer Mark Connelly, WA1ION, as he took an interest in our pre-DXpedition planning.

DXpedition Ground Zero. *Here is the Listening Waters Victorian Farmhouse in Murray Harbour North, on the southeast shore of Prince Edward Island.*

"It's not quite the same as a solo — or maybe a two-person — local deal when at 2 p.m. you say, *Maybe I'll shoot over to the beach for a quick session at sunset,* and that beach is less than an hour's drive from the house," he said.

"If you forget some power cord or a BNC/UHF adapter and then — as you sit at the site in a non-functional DX state — you realize that you will have to bump things to another night, the most you've wasted is maybe a couple of hours driving around.

"If, on the other hand, you've taken days off from work, invested $1,000 or more just to get to (and) from the site, and, even worse, several of your friends who spent just as much to get there are depending on you, then you've really boned up if you left something important at home. *So plan ahead!"*

Strategies via Email: Things to Bring

The packing list was most critical for two DXpedition team members who would be flying in. Airport security, baggage fees, and the potential for lost luggage place limitations on what can be carried without risk. Some of the planning discussions via email exchanges went like this:

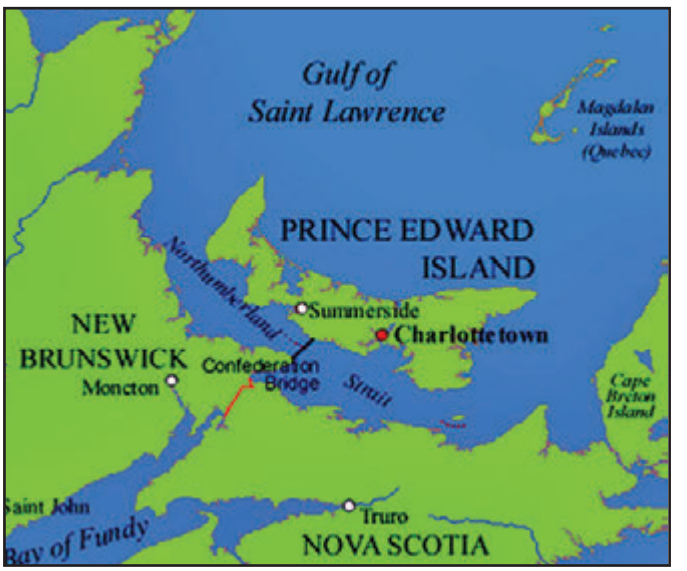
"Counting down the days!" exclaimed Walt. "A question for the group. Perhaps I've forgotten if you've already mentioned what to bring as far as DX equipment. What do you suggest as far as connectors, splitters, antenna switches, coax, etc.? I wouldn't want to come out having forgotten something crucial!"

"I guess, from your perspective, you would want to have at least BNC cables to attach to each splitter and then a switch to select between/among antennas," replied Brent. "I have a Wellbrook active 4-way splitter and I believe Niel does also. I think last year we powered them from one of my 12-V storage batteries so that there is one less switching power supply on the go.

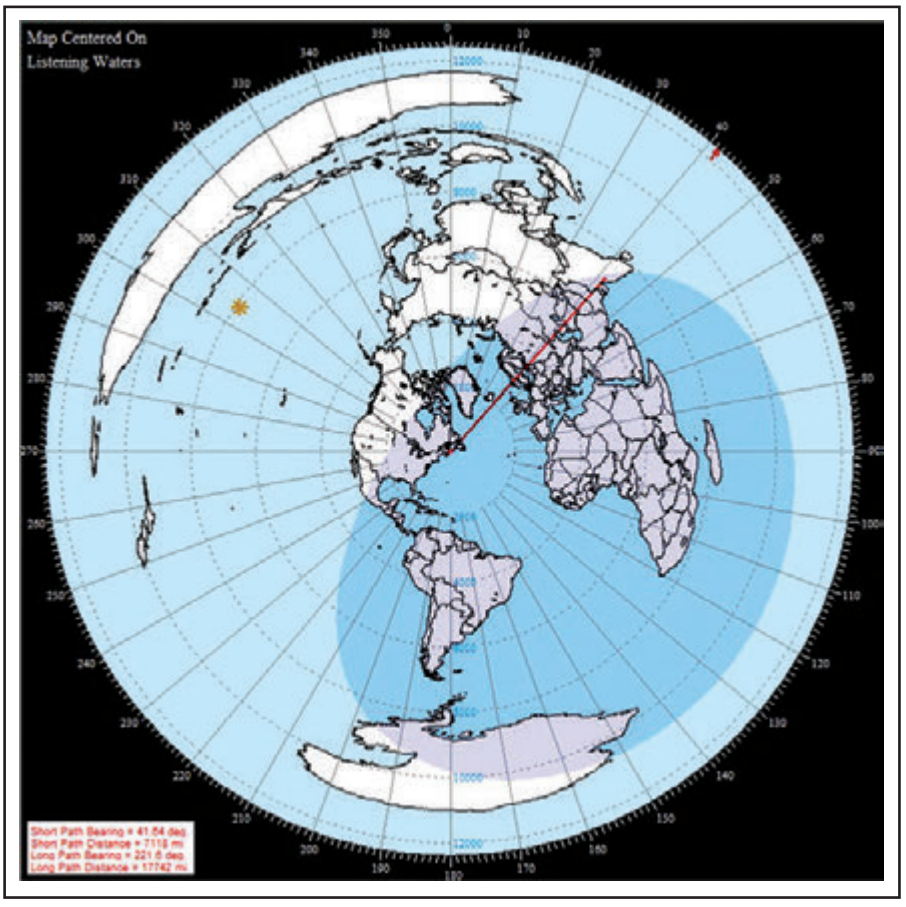
"There will be at least two and as many as 4 or 5 antennas to choose from — and now with Beverage possibilities, maybe even more. One can never have enough splitters and multi-position switches. I'm also going to pack my Quantum Phaser. Bruce has his own custom phasing box and I think Niel just switches on the fly. We'll each — I believe — have access to one feed from each 'community' antenna.

"As for gear power, my battery will be connected to an Anderson PowerPole strip with some empty ports. I also have some spare connectors and a crimper here, which I will ensure I get packed in one of my totes. As for other things, it's never a bad idea to have lots of PL259/BNC adapters and joiners. And, headphones!"

"Believe it or not, Brent, I have forgotten headphones, once!" said Walt. "I'm a BNC fellow. What do the others use? My Alpha Delta switches are PL259, of course, but I have the



Oh, Canada. *Prince Edward Island is part of a vast region in eastern Canada. (Courtesy of Wikimedia Commons)*



Pathfinders. *This AZ mapping of the path to India, a prime DX target, shows a 40-degree bearing from Listening Waters. The Beverage antenna was aimed at 60 degrees toward Europe, midway between India and Africa, which proved to be a good decision.*

adapters for those. My coaxes are all BNC, though. What about lengths? What's the minimum length you'd suggest? Questions, questions!"

"I am a BNC person," remarked Niel. "I pretty much bring everything I can think of that I believe will be useful and that will fit in luggage. I do, indeed, have a Wellbrook active four-way splitter. I think it last saw the light of day last year in Howe Bay."

"Typically Niel and I, and the 'import ringer' DXer set up at a dining room table, and Bruce is adjacent on a table of his own — due to the breadth of his equipment (two laptops, etc.)," said Brent. "So, for our purposes we just need 3-foot lengths for our on-the-table work. We try to port one or two of Bruce's feeds over to our table also, and if there are enough active splitters, we can all share."

"I will have one two-way and one four-way active splitter, both with BNC connections," I replied. "I'm planning on a bi-directional dual Delta aimed toward Japan and Argentina, and a 1,200-foot Beverage aimed toward North Africa. I'll have a couple of 150-foot coax cables, a 75-foot cable, some shorter lengths, twisted pairs, and an assortment of RF matching/isolation transformers for the antenna hook-ups. This week I'm going to build a special transformer for coupling twisted pairs between the two Delta antennas to a coax lead-in."

Aerial Guidance from Above

Antenna planning was facilitated by Google Earth. Satellite imaging allowed for pre-determining antenna layouts and positions. Beverage antenna placement within a 60- to 90-degree bearing would ultimately depend upon agricultural activity in an adjacent field, while there was plenty of room on the 17-acre site for positioning broadband SuperLoop and Delta antennas in any direction.

I was driving my car to the site, so I could bring plenty of wire and coax for antennas. If customs had any questions at the border crossing, I would describe it as *ham radio* gear, something most people understand versus trying to explain a broadcast DXpedition.

"It's always good to bring a back-up receiver," I continued in the email exchange. "There would be nothing worse than traveling a long distance for a DXpedition only to discover that your only receiver crapped out or was damaged during the trip. I will have an Excalibur SDR for the primary receiver, and an RFSpace SDR-IQ for backup. And don't forget to bring an AC power strip so you'll have enough outlets to plug in all your stuff."

The *2013 WRTH*, (World Radio-TV Handbook) and downloads of the latest EMWG (an index of long- and medium-wave stations in Europe <http://www.emwg.info>; PAL (Pacific Asia Log) <http://bit.ly191uqIO>; and MW Offsets (listing of exact frequencies for medium-wave stations) <http://bit.ly/10cLFyz>, were recommended for references. Internet references were to be downloaded to our laptops ahead of time, just in case we didn't have a reliable wireless Internet connection at the DXpedition site.

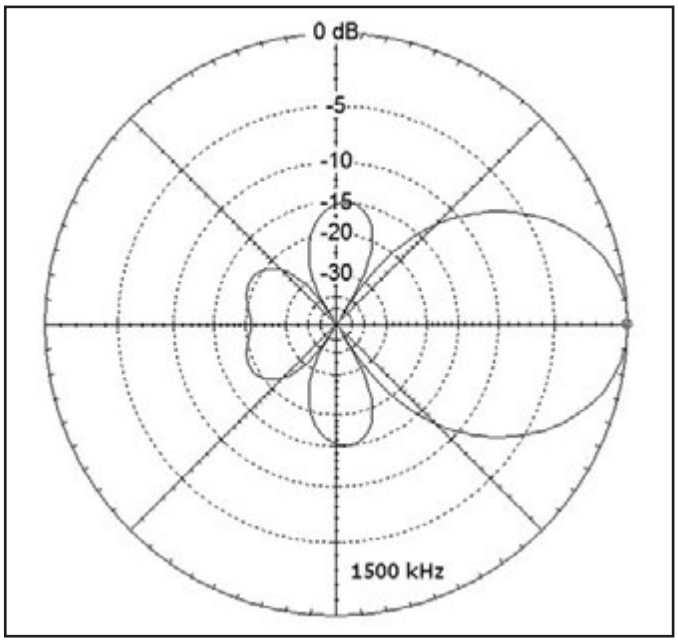
Members of the team also had copies of the NRC AM Radio Log <http://bit.ly/1geTc8M>, and the IRCA Mexican Log <http://www.am-dx.com>. Additionally, we charted sunrise and sunset times along with bearings for various DX target areas. *The DXpedition team was primed and ready to go.*

Savoring the Results

Everyone arrived safely with all their radio gear intact for what was to be an exciting week of AM broadcast DXing. Software-defined-radio receivers ruled the DXpedition, with a Perseus, RFSpace SDR-IQ, and two WiNRADiO Excaliburs capturing all the action, plus an ultralight just for fun.

"Once again, it was a delight to welcome DXers to our little corner of Atlantic Canada," said Brent. "We took full advantage of the 17 acres with a Beverage as well as other antennas. The week was a great time of fellowship, DX, and sharing best practices. While I would have hoped for a more favorable mix of solar-terrestrial indices, the solid captures of India 1071 and Botswana 909 made the whole week worthwhile for me. Grabbing India with my ultralight Sangean DT-400W — inductively coupled to the Beverage wire — was a personal highlight."

"What a treat it was to be totally immersed in fantastic European trans-Atlantic, medium-wave reception, in a lovely location, and very comfortable accommodations," said Walt. "The



A Good Antenna. *EZNEC antenna modeling of a terminated Dual Delta array demonstrates its highly directional capability.*

weather co-operated beautifully, even if the ionosphere didn't always. The thrill of hearing India before 3 p.m. local, and the great camaraderie of seasoned DXers, made the long journey very worthwhile.

"I'm already dreaming of return visits in the years to come. Brent's organizational abilities were much appreciated. Niel's expertise in the MW scene on the East Coast was evident daily, and we all enjoyed his humor. Bruce's technical expertise was very much in evidence, and to watch him rapidly erect Delta loops was something to behold.

"All in all, the total package was a perfect fit for me. Personal highlights in no particular order were: (1) AIR on 1071 daily in the afternoon with the sunshine streaming through the window. (2) LKB/LLE from Norway on 5895 with only 50 watts. (3) The numerous Europirates above 1600 kHz. (4) The magic of having the entire MW band light up starting at 1400 local and continuing throughout the afternoon and late evening, chock block full of European MW stations. Now to just begin to search the TB of Perseus .wav files. Countless hours of memories there, I'm sure!"

"This was another great location," concurred Niel. "Plenty of room for antennas and away from many noise-making devices. Although we did not get any trans-Pacific DX this time, and the Latin Americans were sparse, I'd say we had more than enough from Europe, Africa, and the Middle East to keep us busy from mid-afternoon until it was time to hit the sack. I was especially thrilled to meet Victoria DXer Walt Salmaniw for the first time.

"I had hoped that having him present would be a big help if the trans-Pacific DX began to roll in at sunrise. Still, Walt's presence was handy for identifying the Slavic language broadcasts coming from Europe and also spotting some interesting Europirates above 1600 and shortwave signals, especially the All India Radio outlets."

Assessing the DXpedition's Week

This was my fourth Prince Edward Island DXpedition, after having survived 2009 Priest Pond, 2010 Savage Harbour, and 2012 Howe Bay. Each of these DXpeditions has been a unique experience.

Last year at Howe Bay was most notable for its outstanding tropical reception. *Welcome to Bizarro World* might have been the moniker for this year, as it was the complete opposite of the 2012 experience.

Despite an experimental bi-directional dual Delta array aimed north-south which seemed to be doing the job as it was the only antenna receiving 1220 Brazil, atmospheric conditions weren't cooperating, with otherwise only the usual stuff from Cuba, Puerto Rico, Venezuela, and a couple other regulars received.

The Delta was later terminated to favor south, but still no

Seeing the Sites. *Panmure Lighthouse was among the sights noted as DXpeditioners took time out from their monitoring activities to tour the island.*

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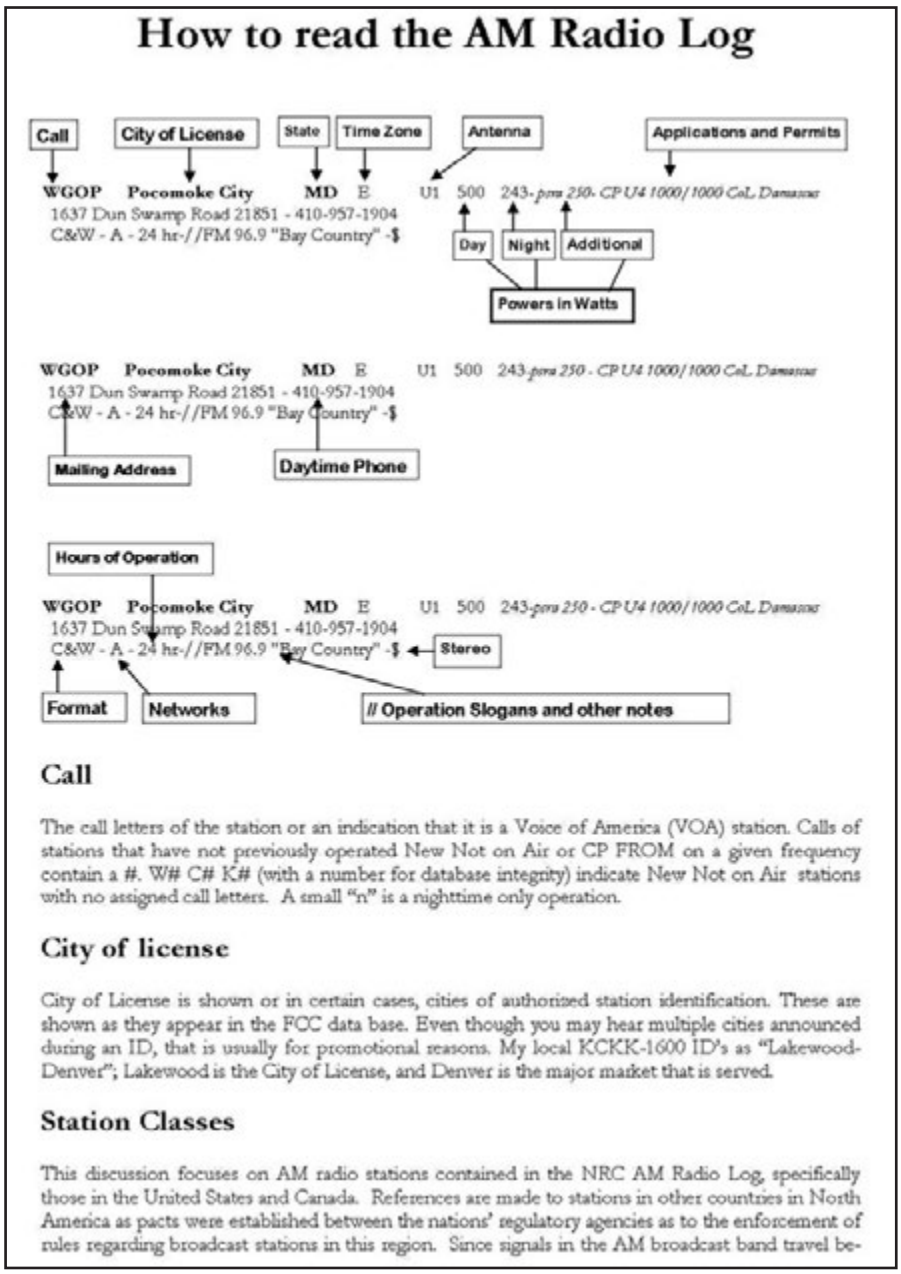
sign of Argentina, Paraguay, and Uruguay. That's OK though, as a trans-Atlantic DX overdose more than made up for it. Check out how we piled on 1071 kHz to hear India. I also enjoyed watching rookie Walt Salmaniw go after those Europirates in the x-band, while stations like "Bluebird" and "Dr. Einstein" provided entertainment between top of the hour RF spectrum captures.

Iran seemed to pop up everywhere, just as noted lately from home base, and the 60-degree Beverage attracted România Actualitati (at 56 degrees) and the U.K. (approximately 60 degrees) like bears to honey. However, there were no bears on this group, as we all collaborated to make this another exceptional DXpedition.

And yes, time does fly when you're having fun. The week flew by so fast that I hardly had enough time to get settled before it was already time to say goodbye.

'Listening Waters' Loggings

For a complete rundown of the DXpedition's loggings, visit: <http://www.CQPlusMediumWaves.blogspot..com>.



How to read the AM Radio Log

Call | City of License | State | Time Zone | Antenna | Applications and Permits

WGOP Pocomoke City MD E U1 500 243-*pwra 250- CP U4 1000/1000 CoL Damascus*
1637 Dun Swamp Road 21851 - 410-957-1904
C&W - A - 24 hr-//FM 96.9 "Bay Country" -$

Day | Night | Additional

Powers in Watts

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C&W - A - 24 hr-//FM 96.9 "Bay Country" -$

Mailing Address | Daytime Phone

Hours of Operation

WGOP Pocomoke City MD E U1 500 243-*pwra 250 - CP U4 1000/1000 CoL Damascus*
1637 Dun Swamp Road 21851 - 410-957-1904
C&W - A - 24 hr-//FM 96.9 "Bay Country" -$ | Stereo

Format | Networks | // Operation Slogans and other notes

Call

The call letters of the station or an indication that it is a Voice of America (VOA) station. Calls of stations that have not previously operated New Not on Air or CP FROM on a given frequency contain a #. W# C# K# (with a number for database integrity) indicate New Not on Air stations with no assigned call letters. A small "n" is a nighttime only operation.

City of license

City of License is shown or in certain cases, cities of authorized station identification. These are shown as they appear in the FCC data base. Even though you may hear multiple cities announced during an ID, that is usually for promotional reasons. My local KCKK-1600 ID's as "Lakewood-Denver"; Lakewood is the City of License, and Denver is the major market that is served.

Station Classes

This discussion focuses on AM radio stations contained in the NRC AM Radio Log, specifically those in the United States and Canada. References are made to stations in other countries in North America as pacts were established between the nations' regulatory agencies as to the enforcement of rules regarding broadcast stations in this region. Since signals in the AM broadcast band travel be-

DXpedition Reference. *Here's the legend page of the National Radio Club AM Radio Log. The PEI team made sure to download pertinent information in case there was no Internet access at the Listening Waters site. (Internet screen grab <http://bit.ly/1geTc8M>)*

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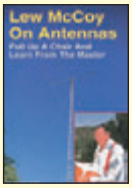

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BY CORY GB SICKLES, WPC2CS

Freeband: Intervention in a 'Love-Hate' Relationship

Now, stick with me. This setup is a bit long, but you'll soon see the necessity.

In February, we pause the bland, windy chill of winter to decorate with hearts, roses, and other signs of Valentine's Day. This increasingly commercial holiday is a boon for florists and chocolatiers, as well as restaurant owners.

Many a romantic proposal is executed on February 14, either in a public setting or a more private one.

As relationships move further along, though, we may find ourselves in something of a "love-hate" relationship with a *significant other* — a not-too-perfect mix of what we like and what we don't. Some call that "a typical marriage."

Well, this month I'm not thinking about a person — rather, *an activity*. It's one that I don't like, but have also found something about it to admire. It's much like consideration for Attila the Hun: If your village was one of those pillaged, you might find it difficult to admire his management style.

Outside the Lines

While Citizens Band operation beyond the designated channels has been around for some time, advances in technology have made the pursuit easier than ever.

*<wa3uvv@gmail.com>

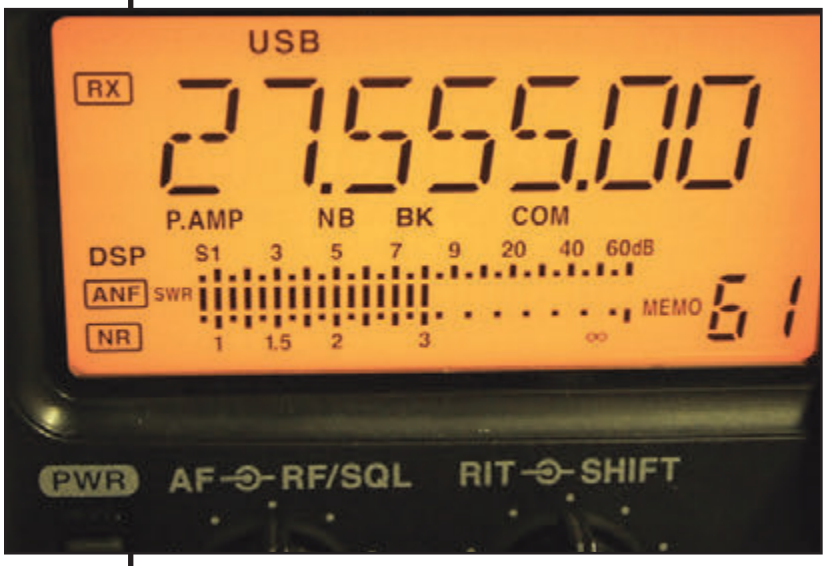

Photo A. *A popular Freeband meeting place — albeit illegal — is 27.555 MHz — not a frequency allocated for Citizen Band operation. (Courtesy of WPC2CS)*

Back in the days of the crystal-mixing-ways of generating 23 channels, some folks worked out ways to access channels 22A, 22B, 23A, and 23B — now known as 24, 25, 26, and 27, respectively, **Table 1**.

Still others, with 5- or 6-channel rigs that used a pair of crystals for each channel, occupied the radio control allocations of 3A, 7A, 11A, 15A, and 19A for communications that was a bit quieter and more secluded — while possibly sending the occasional radio-control (RC) model car to intermittently take off or shudder to a halt.

Still others found ways to incorporate variable frequency oscillator (VFO) technology — also known as a "slider" — to travel pretty much all over the 27-MHz region.

At the same time, others just re-configured amateur radio transceivers of the day to cover this unauthorized spectrum — with 100 to 500 watts of power; much more than allowed.

And With Its Own Name

The term identifying this clandestine slice of spectrum between CB channels is Freeband, occupied by Freebanders who are Freebanding.

All of this activity was — and remains — illegal. Thus, it was with a grimace and frown when I discovered a Freeband watering hole — 27.555 MHz, **Photo A**. This is apparently used by an international group of two-way communications enthusiasts as a calling channel.

That's much like CB Channel 11 was *supposed* to be:

- Make a call for a contact
- When someone answers, move to another channel to talk
- You leave the calling channel free when you move to another channel to chat

With '555, you place a call and let others know you're listening 10 kHz up, 20 kHz down or wherever is presumably clear.

All activity seems to be via SSB, leaving AM behind in these activities.

A Redeeming Value?

While I seriously frown on the whole idea of pirate operations on frequencies allocated to other services, I have to admit that in the case of '555, the channel runs in a relatively polite and professional manner — not unlike amateur radio. (***UNDERSCORE:*** *I'm not condoning this illegal activity. I'm merely marveling at its existence and*

dutifully reporting about it. ***VISUAL:*** *Right now, I'm flailing myself with wet rutabaga leaves just for drawing the Freeband-amateur radio comparison. – WPC2CS.)*

QSL cards are typically exchanged and my limited research shows there even seems to be a loosely run awards program for counties, states, provinces, countries worked, and so on.

As radio amateurs have used the Internet to enhance their interests with sites such as <http://www.QRZ.com>, so have the 11-meter Freebanders with <http://www.QRZ11.com>, **Photo B**. The 27.555-MHz Freeband crowd even has a Facebook page! Several, in fact! *(**VISIT:** For example, the 27.555-Mhz International Frequency DX page at <http://on.fb.me/1dwiJge>,* ***Photo C****. – WPC2CS.)*

A glance through some of the group members' pictures show impressive-looking stations.

27.555 Operators' Gear

What equipment are these Freebanders using? Most seem to have re-tasked amateur radio HF gear and more than a few amplifiers — many of which could produce more power than

Photo B. *The highly-organized Freeband community has a website modeled somewhat on the amateur radio community's QRZ.com. For Freebanders, it's the QRZ11 Forum at <http://www.QRZ11.com>) (Internet screen grab)*

Photo C. *Among several Facebook pages established by, and devoted to Freebanders is the 27.555-Mhz International Frequency DX page at <http://on.fb.me/1dwiJge>. (Internet screen grab)*

even amateurs are allowed. Others have more modest stations, based on "import" radios that allow for multiple blocks of 40 channels, which enable Freeband activities.

Such manufacturers and sellers in the United States skirt the law by marketing them as "10-meter" ham rigs. Trust me, hams aren't buying these, as they are too limited for our purposes and desires. Just cut the magic yellow, blue, or green wire inside (instructions, of course, on the Internet) and you are "good to go" with joining the Freebanders.

FCC Enforcement (?)

While I *do read* about cases where someone is investigated and fined by the Federal Communications Commission — *Huzzah!* — for excessive power and interference, I sadly don't see much, *if any,* in the way of stories about enforcement toward "out of band" activity. Occasionally, a company like ePower Amps will be fined for selling illegal amplifiers (a.k.a. CW transmitters), but I was seeing Facebook ads for its products months ahead of the crackdown. Who knows how many they sold before that? Plus, since the inventory was apparently not confiscated, who knows how they eventually may have "disposed" of the inventory? *(**IN DEPTH:** Read an FCC Citation and Order for "Illegal Marketing of Unauthorized Radio Frequency Devices" against ePower Amps at <http://bit.ly/1i9lxlp>. – WPC2CS.)*

I guess it's a matter of priorities. There's more illegal activity to go around than the Commission can address. Broadcasters and other business owners are easier targets — high profile, with serious bank accounts. If the harmonics of some poorly-made amplifier interfere with aviation or public service, then it becomes a higher priority. If not, it seems to be ignored.

Factor in all of the other countries and agencies involved and I imagine the task of enforcement and "putting the toothpaste back in the tube" becomes incredibly daunting. No doubt, that's what the Freebanders are counting on, too.

Simply Put ...

For me, *CB is CB* and *ham radio is ham radio.* I have all sorts of transceivers that could be modified and used for such clandestine activity and it would technically be easy to join the crowd on 27.555 (and 26.985) MHz.

But for more than 45 years, I've chosen to keep the two separate and enjoy each for what it is. *So have most of us.*

While I am duly impressed at the scale and sophistication that this activity has risen to, I have to ask why these folks don't just channel their energies — and money — into amateur radio?

An Invitation to Freebanders

If you are a Freebander and would like to "clean up your act" to experience a broader world of *legal* communications, I think the amateur radio community would welcome you. If the thrill of working outside of the law is the most important thing, then please stay where you are.

27-MHz Channels and Frequencies

Designators in this listing are:
** Class C Radio Control*
*** Original Designator*

CHANNEL	FREQUENCY	CHANNEL	FREQUENCY
1	26.965 MHz	19A*	27.195 MHz
2	26.975 MHz	20	27.205 MHz
3	26.985 MHz	21	27.215 MHz
3A*	26.995 MHz	22	27.225 MHz
4	27.005 MHz	23	27.255 MHz
5	27.015 MHz	24	27.235 MHz (22A)**
6	27.025 MHz	25	27.245 MHz (22B)**
7	27.035 MHz	26	27.265 MHz (23A)**
7A*	27.045 MHz	27	27.275 MHz (23B)**
8	27.055 MHz	28	27.285 MHz
9	27.065 MHz	29	27.295 MHz
10	27.075 MHz	30	27.305 MHz
11	27.085 MHz	31	27.315 MHz
11A*	27.095 MHz	32	27.325 MHz
12	27.105 MHz	33	27.335 MHz
13	27.115 MHz	34	27.345 MHz
14	27.125 MHz	35	27.355 MHz
15	27.135 MHz	36	27.365 MHz
15A*	27.145 MHz	37	27.375 MHz
16	27.155 MHz	38	27.385 MHz
17	27.165 MHz	39	27.395 MHz
18	27.175 MHz	40	27.405 MHz
19	27.185 MHz		

Table 1.

The Need for a Smoothed Solar Index — Accurate Propagation Predictions, Part II

BY CARL LUETZELSCHWAB, K9LA

In the first of this two-part series, I discussed why we need a smoothed solar index — to best characterize a solar cycle. This month's column addresses a second reason why we need a smoothed solar index — to achieve the best accuracy with propagation predictions. *(**NOTE:** See Part I on the CQ Plus Internet landing page at <http://www.CQPlusPracticalPropagation.blogspot.com>. – KI6SN)*

To analyze this accuracy issue, we'll focus on the *CQ World Wide SSB* contest, which took place at the end of October last year. By the way, the higher bands were excellent.

We'll use ionosonde data for the one-month period centered on October 26 and 27 — the weekend of the contest. We'll use the Boulder (Colorado) ionosonde from October 12 through November 10 at 1900 UTC as it had tabular data for both October and November 2013. **Figure 1** presents this data, along with the 10.7-cm solar flux for the same period.

There is a lot of information in **Figure 1**.

- The solid orange curve is the daily MUF (maximum useable frequency) over the Boulder ionosonde assuming it's the midpoint of a 3,000-km (1,864 miles) F_2 region hop. Note that it varied from a low of 29 MHz on October 12 to a high of 37 MHz on November 8. There were many ups and downs throughout the one-month period.
- The dotted orange curve is a linear trend line for the MUF and indicates that the MUF had an overall increasing trend during the period.
- The solid blue curve is the daily 10.7-cm solar flux. It started at a value of 128, dipped to a low of 125 on October 14 and 15, peaked at a high of 167 on October 27, decreased to around 140, and then increased at the end of the period.
- The dotted blue curve is a linear trend line for the 10.7-cm solar flux, and indicates that the 10.7-cm solar flux also had an overall increasing trend during the period.

We can draw two important conclusions from this data. The first is that the MUF and solar flux do not appear to be very well correlated on a daily basis. This is best seen from October 21 to October 30, when the solar flux reached its highest values but the MUF was up and down significantly. In fact, the highest solar flux on October 27 resulted in one of the lowest MUFs during the entire period.

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A better way to see the correlation between the daily solar flux and the daily MUF is through a scatter plot. Each daily solar flux value and daily MUF value is plotted and then a trend line is added to determine the correlation factor R-squared.

- An R-squared value of 1.00 means perfect correlation — all the data points would be on the trend line.
- An R-squared of 0.00 means no correlation — all the data points would be scattered widely about the trend line.

Figure 2 is the scatter plot for the data in **Figure 1**.

As expected, the extremely low value of R-squared (highlighted in red) tells us the correlation between the daily solar flux and the daily MUF is poor. For example, for a solar flux of 128, the MUF could have been 29 MHz or 34.5 MHz. In other words, a daily solar flux value does not map to a unique daily MUF.

The second conclusion is due to the fact that the solar flux and MUF trend lines are parallel, which tells us that solar flux and MUF follow a linear relationship. But this relationship is on a long-term basis — not on a daily basis. As can be seen, the higher the solar flux, the higher the MUF. This should not be new news. The fact that this does not apply on a daily basis may be new to you.

Why doesn't the daily MUF follow the daily solar flux? The reason is that although solar radiation certainly instigates the ionization process, ultimately the ionization at any given location is also

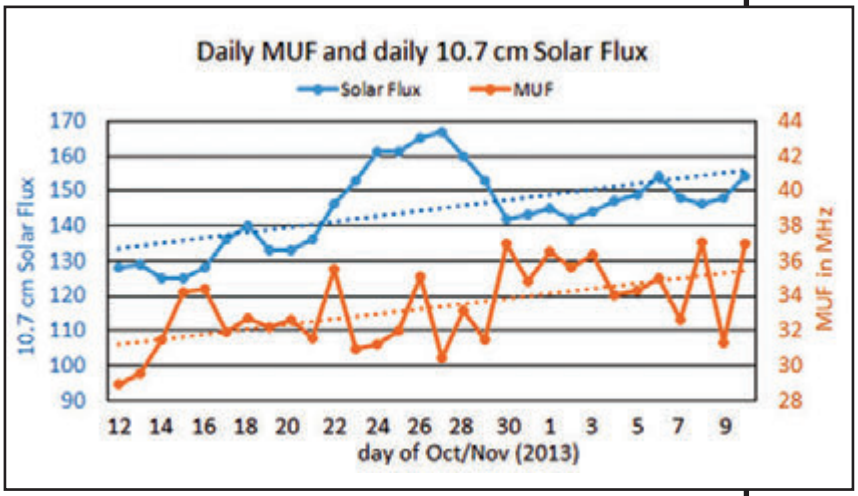


Figure 1. *Solar Flux and Boulder Data*

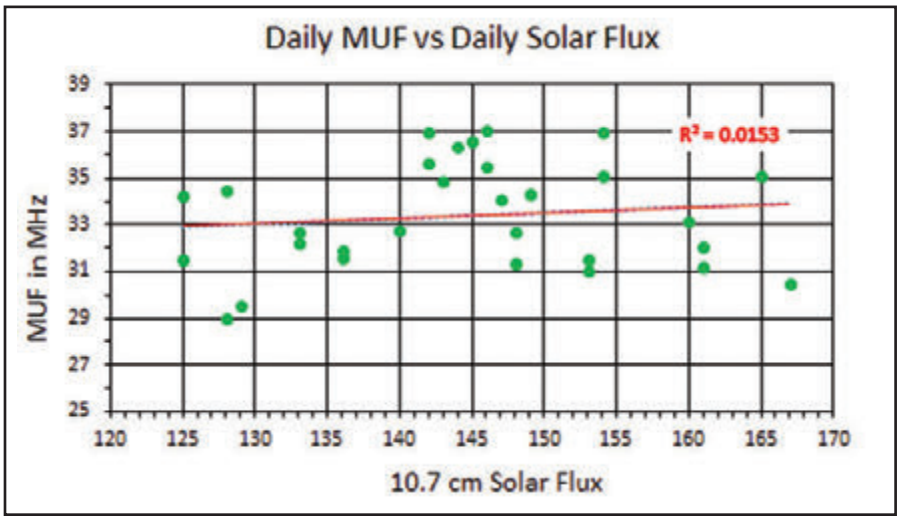


Figure 2. *Scatter Plot*

dependent on geomagnetic field activity and events in the lower atmosphere coupling up to the ionosphere.

*(**IN DEPTH:** See my August 2004 WorldRadio column titled "The Day-to-Day Variability of the Ionosphere" for more details on this issue. Visit <http://www.k9la.us>. Look under the General link. Included in the piece are scatter plots showing the poor correlation between the daily 10.7-cm solar flux and the daily MUF over the Millstone Hill (Massachusetts) ionosonde. – K9LA.)*

If the daily correlation is poor, how was a model of the ionosphere developed for propagation prediction purposes? When scientists discovered that the correlation was poor between daily values, they moved on to a longer-term correlation — which turned out to be between a smoothed solar index and monthly median parameters of the ionosphere. What this means is our model of the ionosphere is a monthly median model. This inherently says our propagation predictions are probabilities over a month's time frame.

To better understand the improved correlation with a smoothed solar index, let's look at some scatter plots that compare the daily correlation to the smoothed correlation. We'll do this for the Boulder ionosonde (but at a different date than the data in **Figure 1**) and the Canberra (Australia) ionosonde. **Figure 3** shows this data.

At both ionosondes, the daily solar flux versus daily MUF results in poor correlation. But the correlation between the smoothed solar flux and the monthly median MUF is extremely high.

What this means is a smoothed solar flux value maps much better to a unique monthly median MUF. This is what is needed for accurate propagation predictions, and is the second reason why we need a smoothed solar index.

Next month's column will review smoothed solar indices, monthly median ionospheric parameters and the interpretation of propagation predictions.

73, Carl, K9LA

NW7US February HF-VHF Outlook
For Tomas Hood, NW7US's SWL and VHF propagation outlook, visit <http://www.CQPlusPropagation.blogspot.com>.

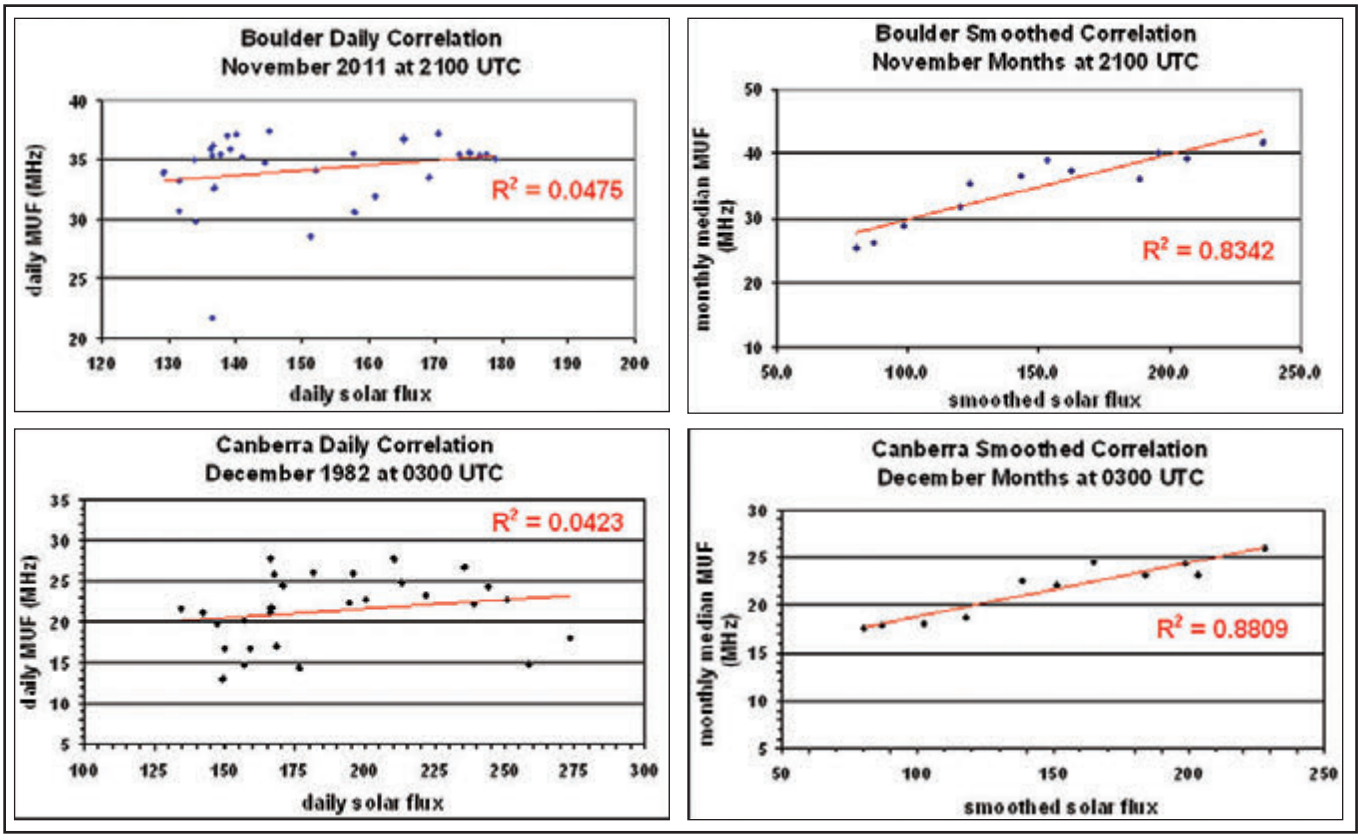


Figure 3. *Daily versus Smoothed Correlation*

Optimum Working Frequencies (MHz) - For February 2014 - Flux = 132, Created by NW7US

UTC TO/FROM US WEST COAST	00	01	02	03	04	05	06	07	08	09	10	11	12	13	14	15	16	17	18	19	20	21	22	23
CARIBBEAN	24	22	17	14	13	13	12	12	11	11	11	11	11	11	18	22	24	25	26	26	27	26	26	25
NORTHERN SOUTH AMERICA	33	31	27	20	19	18	17	16	16	15	15	15	14	14	20	28	30	32	34	35	36	36	35	34
CENTRAL SOUTH AMERICA	33	30	26	19	18	17	16	16	15	15	15	14	14	14	23	29	31	32	33	34	35	35	35	34
SOUTHERN SOUTH AMERICA	35	33	30	26	22	20	19	18	17	16	16	15	15	15	14	26	30	32	33	34	35	36	36	36
WESTERN EUROPE	11	10	10	10	10	10	10	10	10	10	10	10	10	10	13	17	18	19	18	17	16	12	11	11
EASTERN EUROPE	10	10	10	10	10	9	9	10	10	10	10	10	10	10	10	12	12	12	11	11	10	10	10	10
EASTERN NORTH AMERICA	26	24	20	15	15	14	13	13	13	12	12	12	12	12	20	24	26	28	28	29	29	29	28	27
CENTRAL NORTH AMERICA	15	14	13	10	8	8	8	7	7	7	7	7	7	7	7	12	13	15	15	16	16	16	16	16
WESTERN NORTH AMERICA	8	8	7	6	4	4	4	4	4	4	3	3	3	3	3	4	6	7	8	8	8	8	8	8
SOUTHERN NORTH AMERICA	25	24	21	16	14	13	13	12	12	12	11	11	11	11	11	20	23	25	26	27	27	27	27	26
HAWAII	23	23	22	21	19	16	13	12	11	11	10	10	10	10	10	9	9	16	19	21	22	23	23	24
NORTHERN AFRICA	11	11	10	10	10	10	10	10	10	10	10	10	10	10	15	18	19	20	21	19	13	12	12	11
CENTRAL AFRICA	14	13	13	12	11	11	11	10	10	10	10	10	10	10	14	17	19	20	20	18	16	15	15	14
SOUTH AFRICA	24	22	17	14	14	13	13	12	12	12	12	12	12	11	20	23	25	26	27	27	27	27	26	26
MIDDLE EAST	10	10	10	10	10	11	11	10	10	10	10	10	10	10	10	15	17	18	13	12	12	11	11	11
JAPAN	22	22	21	21	19	17	12	12	11	11	11	10	10	10	10	10	10	10	10	10	10	17	20	21
CENTRAL ASIA	22	22	21	20	19	17	12	12	11	11	11	10	10	10	10	10	10	13	13	12	12	12	17	22
INDIA	17	17	17	16	13	11	11	10	10	10	10	10	10	10	9	9	10	10	10	9	9	11	14	16
THAILAND	21	21	21	20	18	16	12	12	11	11	11	10	10	10	10	10	10	14	14	14	13	13	13	12
AUSTRALIA	29	30	32	32	29	25	19	18	17	16	16	15	15	15	14	14	14	19	18	17	18	22	25	27
CHINA	19	21	20	19	18	15	12	12	11	11	10	10	10	10	10	10	10	10	10	10	10	10	10	16
SOUTH PACIFIC	33	34	34	32	29	23	21	19	18	17	17	16	15	15	15	15	18	18	18	22	25	28	30	32

UTC TO/FROM US MIDWEST	00	01	02	03	04	05	06	07	08	09	10	11	12	13	14	15	16	17	18	19	20	21	22	23
CARIBBEAN	26	23	19	18	17	16	15	14	14	13	13	13	12	19	24	27	29	30	31	31	31	30	30	28
NORTHERN SOUTH AMERICA	30	27	23	21	20	18	17	16	15	15	14	14	14	18	23	26	29	30	32	33	33	33	33	32
CENTRAL SOUTH AMERICA	31	28	23	21	20	19	18	17	16	16	15	15	15	24	27	29	31	32	33	34	35	35	35	34
SOUTHERN SOUTH AMERICA	34	32	27	25	23	21	20	19	18	17	16	16	15	16	26	28	30	32	33	34	35	36	36	36
WESTERN EUROPE	11	10	10	10	10	10	10	10	10	10	10	10	10	16	19	20	20	20	20	19	17	14	12	11
EASTERN EUROPE	10	10	10	10	10	10	10	10	10	10	10	10	10	14	16	16	16	16	15	14	11	11	10	10
EASTERN NORTH AMERICA	18	16	11	11	10	10	10	9	9	9	9	9	9	13	17	19	20	21	21	21	21	21	20	20
CENTRAL NORTH AMERICA	9	8	7	5	5	5	4	4	4	4	4	4	4	4	7	8	9	9	9	10	10	10	9	9
WESTERN NORTH AMERICA	15	14	13	10	9	8	8	8	7	7	7	7	7	7	7	12	14	15	16	16	16	16	16	16
SOUTHERN NORTH AMERICA	17	16	14	10	10	9	9	9	8	8	8	8	8	8	13	16	17	18	19	19	19	19	19	18
HAWAII	27	26	24	22	16	14	13	13	12	12	12	11	11	11	11	11	11	20	23	26	27	28	28	27
NORTHERN AFRICA	14	13	12	12	11	11	10	10	10	10	10	10	12	18	20	21	22	23	23	23	20	16	15	14
CENTRAL AFRICA	14	13	12	12	11	11	10	10	10	10	10	10	12	18	20	21	22	22	23	21	18	17	16	15
SOUTH AFRICA	24	19	18	17	16	16	15	15	15	14	14	14	19	27	31	33	34	35	35	35	34	32	29	26
MIDDLE EAST	10	10	10	10	10	10	10	10	10	10	10	10	10	16	19	20	21	19	16	13	12	11	11	11
JAPAN	21	21	19	17	13	12	11	11	11	10	10	10	10	10	10	10	10	10	10	10	10	16	19	21
CENTRAL ASIA	21	20	19	17	12	12	11	11	11	10	10	10	10	10	10	12	13	13	13	12	12	12	13	22
INDIA	13	15	14	12	11	11	11	10	10	10	10	10	10	10	14	17	16	14	11	11	10	10	10	10
THAILAND	20	19	18	15	12	11	11	11	10	10	10	10	10	10	10	14	16	15	14	14	13	13	13	12
AUSTRALIA	29	31	31	28	21	18	17	17	16	15	15	15	15	14	14	16	20	19	18	17	19	22	25	27
CHINA	18	19	18	15	12	11	11	11	10	10	10	10	10	10	10	11	11	10	10	10	10	10	10	14
SOUTH PACIFIC	34	34	32	28	22	21	19	18	17	17	16	16	15	15	15	20	19	18	20	24	27	29	31	33

UTC TO/FROM US EAST COAST	00	01	02	03	04	05	06	07	08	09	10	11	12	13	14	15	16	17	18	19	20	21	22	23
CARIBBEAN	21	17	16	15	14	13	12	12	11	11	11	10	13	18	21	23	24	25	25	25	25	25	24	23
NORTHERN SOUTH AMERICA	27	25	22	20	19	17	16	15	14	14	13	13	14	19	22	25	26	28	29	29	30	30	29	28
CENTRAL SOUTH AMERICA	31	28	25	23	22	20	19	18	17	16	16	15	22	25	28	29	31	32	33	34	35	35	35	33
SOUTHERN SOUTH AMERICA	33	30	28	25	23	22	20	19	18	17	16	16	19	24	27	29	31	32	33	34	35	36	36	35
WESTERN EUROPE	11	10	10	10	10	9	9	9	9	9	9	14	18	19	21	21	21	20	20	19	17	15	11	11
EASTERN EUROPE	10	10	10	10	10	10	10	10	10	10	10	12	17	19	19	19	19	18	17	16	13	11	11	11
EASTERN NORTH AMERICA	8	6	5	5	5	5	4	4	4	4	4	4	5	8	9	10	10	10	10	10	10	10	10	9
CENTRAL NORTH AMERICA	19	17	12	11	11	10	10	10	10	9	9	9	9	14	18	20	21	22	22	23	22	22	21	21
WESTERN NORTH AMERICA	26	24	20	15	15	14	14	13	13	13	12	12	12	12	21	24	26	28	29	29	29	29	29	28
SOUTHERN NORTH AMERICA	21	18	14	13	12	12	11	11	10	10	10	10	10	15	19	21	23	24	24	24	24	24	23	22
HAWAII	27	25	21	16	15	14	14	13	13	13	12	12	12	12	13	13	12	22	26	28	30	30	29	28
NORTHERN AFRICA	14	13	13	13	12	12	12	12	12	12	12	21	24	27	28	29	29	29	27	25	21	16	15	14
CENTRAL AFRICA	15	15	14	13	13	13	12	12	12	12	12	21	25	27	28	29	30	29	27	23	20	18	17	16
SOUTH AFRICA	23	22	20	19	18	17	16	16	15	15	15	23	29	32	34	35	35	35	35	35	34	32	29	27
MIDDLE EAST	13	12	11	11	11	10	10	10	10	10	10	16	19	21	22	23	23	24	22	16	15	14	14	13
JAPAN	19	17	13	12	11	11	11	10	10	10	10	10	10	11	11	10	10	10	10	10	10	14	18	20
CENTRAL ASIA	18	16	12	12	11	11	11	10	10	10	10	10	10	15	14	14	13	13	13	12	12	12	12	20
INDIA	10	10	10	10	10	10	10	10	10	10	10	10	14	18	18	17	17	16	15	12	11	11	10	10
THAILAND	15	12	12	11	11	11	10	10	10	10	10	10	12	17	18	17	16	15	14	14	13	13	13	12
AUSTRALIA	30	30	25	19	18	17	17	16	15	15	15	15	14	14	23	21	20	19	18	17	19	23	26	28
CHINA	16	13	12	11	11	11	10	10	10	10	10	10	10	12	12	11	11	10	10	10	10	10	10	10
SOUTH PACIFIC	34	31	27	24	22	20	19	18	17	17	16	15	15	16	21	20	18	18	23	26	29	31	33	34

On an Electromagnetic Wing and a Snowy Winter Evening's Prayer

BY SHANNON HUNIWELL, WPC2HUN

Metaphorically, AM skywave signals are a bit like little joys in life. They're capriciously fleeting, though sometimes manage to have the power for appearing in the most unpredictable places. If they're long enough to be recognized by name, such a definitive callsign identified greeting almost always brings its recipient cheer.

My father has a favorite winter DX story that exemplifies this radio reception/life comparison. Dad made me a bit suspicious by prefacing the following tale with an uncharacteristic admission that he can't recall who told it to him, exactly where it occurred, or to what station the saga relates. After tightly closing his eyes and rubbing his temples, he was able to guesstimate that the story's listener was navigating some treacherously ice-covered road in northern Michigan and inadvertently snagged a 250 watter "approximately in West Virginia." Even so, he is willing to swear that "the episode took place just before midnight on a snowy winter night ... probably in the early 1970s."

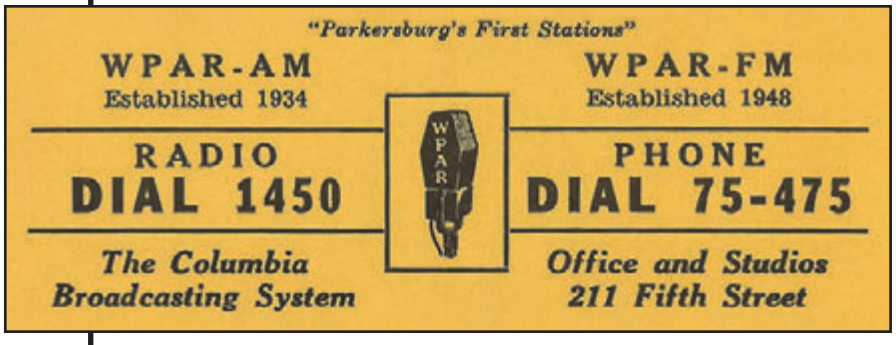
Dad, Are You Sort Of Reasonably Sure That This One Is At Least *Semi-Authentic?*

According to Huniwell legend, the driver of a 1963 Studebaker Champ pickup truck — how my father remembers that automotive detail above broadcast facts, always made me wonder, but anyway — this fellow had promised his elderly mother that he would come home for a visit sometime during the holidays. The old homestead was located near the top of the Upper Michigan Peninsula, a bit of a hike from the busy little St. Louis-area tool and die manufacturing shop where he served as manager. Crunch work had required him to oversee several priority projects that left him and a couple of loyal employees occupied all but part of Thanksgiving and Christmas day. Happily, their tenacity enabled the company to meet its customers' robust orders, and the firm's delighted owner rewarded his staff with a week of vacation starting on December 31st of whatever year Dad couldn't envision. The Studebaker driver — we'll call him, Joe — packed a few clothes, boxed a sample of the parts that had kept him so tied up, tossed the stuff on the front seat of the truck, and headed north to surprise his mom. Conveniently, the Studebaker was allegedly equipped with a "pretty darn good AM radio," an electronic feature that the fellow used as a constant roadway companion throughout the journey's many miles and hours.

My poetic license will have to kick-in here, as I conjecture that when the traveler encountered the icy highway somewhere around Routes 131 and 66 in Kalkaska County, Michigan. Local radio reception way up there and near the witching hour was probably none too great. Arguably, Joe was likely dialing big city skip or maybe a thread of 50-kilowatt ground wave from Chicago or Detroit.

Whatever was tickling the tuner, though, left suddenly when the Studebaker hit a patch of black ice that sent it spinning across the opposite lane. This rotation rapidly propelled Joe's luggage from the seat to the dashboard. A corner of the box in which the sample parts were packaged catapulted against the radio's tuning buttons before dumping its contents onto the passenger floor. Its crescendo coincided with the lengthy blast of an 18-wheeler's air horn from an oncoming tanker that had seemingly come out of the snowy nowhere. Joe braced himself for the impact and flash, but it never came. Instead, he found himself shaken-up and tightly gripping the wheel of his Studebaker, a trusty truck that was still very much intact, though, about a dozen feet off of the road and pointing south. Its engine had stalled, but when Joe regained his composure, he became conscious of a strong signal pouring through the radio.

e-mail: <Melodyfm@yahoo.com>



It's any reader's guess which West Virginia AM was fortuitously snagged by our first story's main character, Joe, while hundreds of miles away from "The Mountain State." Because my father, Sid Huniwell, was reticent about recalling the subject station's identity, I always suspected it might actually have been 50,000-watt country music giant WWVA 1170. "Absolutely not that flamethrower!" Dad would protest. "Besides," he claimed, "Joe's catch was definitely some tiny West Virginia outlet on a local channel." As such a broadcast facility arguably occupied some Class IV, "graveyard," or local channel, it might have been Parkersburg, WPAR 1450 — noted on this advertising card. Then again, 1240 WDNE, Elkins; WMON 1340, Montgomery, WV; or maybe Bluefield's WKOY 1240 could have been among the mighty little 250-watt signals that skipped to the tale's upper Michigan reception point.

In a decided drawl, the bucolic announcer was wishing whomever "might be listening" a happy New Year. The DJ hoped that the occasion would provide folks in his typically tiny audience opportunity for "visiting and cheer." And then the radio host introspectively mused about having lost his wife during the previous winter and that his daughter lived in California and they hadn't seen each other since the funeral. "But don't mind me," he said, "I should be glad that the Good Lord saw fit to let me stay at the microphone for another New Year's Day." With that, the announcer noted it was officially midnight, invited listeners back at 6 a.m., when he would be filling in for the regular morning DJ, and then read the sign-off copy.

Joe was incredibly surprised to hear that the station hailed from a little town in West Virginia and used only a quarter kilowatt after sunset. These facts prompted him to wonder what the announcer there might think if he were to know that Joe had received the fellow's heartfelt New Year's Eve message in a shallow ditch aside a slippery Michigan highway. Joe switched on his truck's dome light, fished a pencil from the glove compartment, and jotted the station's call letters and city of license on the cardboard box he'd retrieved from the floor. He hadn't even finished the fourth letter when the scratchy, wowing LP track of Auld Lang Syne, on a warped Guy Lombardo record that the old announcer had selected as his swan song, evaporated into the ether.

When one thinks of vintage pickups, the long-defunct Studebaker marque seldom comes to mind. While it is doubtful that the formally-dressed admirers in this 1960 ad realistically represent eager truck buyers, the pictured vehicle did attract enough attention from those wanting to show Ford and Chevy owners that some folks didn't care about following the commonplace crowd. Whether Joe's edition of the Champ was equipped with six or eight cylinders is unknown, though his did sport an optional AM push-button radio.

Mail Call!

This story's epilogue simply includes that Joe offered a thankful prayer to God for having spared him, for the Studebaker getting right back on the road, and then making it to his mother's house a couple of hours later, as well a note he penned to *The DJ On Duty New Year's Eve*, in care of the station's calls and locale.

Nearly a year later, a reply arrived. At first, Joe wondered why he was getting correspondence from a radio station in West Virginia, but the message inside quickly jogged his memory and warmed his heart. Again, I'm using poetic license here, although Sid Huniwell, my radio nut father who first told me this saga, claims it's "spot on and nearly word-for-word" what was written on Joe's correspondence:

Greetings from West Virginia,

First off, I want to officially confirm that you did, indeed, hear our signal all the way up in Michigan last December 31st! That's quite a haul for 250 watts part way across this great country. Our contract engineer thinks it is the farthest reception report we ever got and figures that conditions were just right for the skywave skip, especially since there are probably 30 stations between here and there on our local or Class IV frequency. So, it's a feat for us, too! Any DX'er (or long-distance radio listening buff) would be proud to post verification of such a catch in his radio shack (or on his pick-up truck dashboard, as the case may be)!

What I most want to tell you, however, is that your note made our old Bill Shumaker smile. He was the fellow you

My radio buff father emailed me this eBay seller's image of an early 1960s Studebaker radio. He wanted me to be sure to note the tiny CONELRAD Civil Defense triangle symbols at 640 and 1240 kilocycles. I was most interested, however, in the AM set's five plastic push buttons like the one that got rammed with flying cargo and ended up tuning some very good DX.

heard last New Year's Eve. Word of your long-distance reception even made a mention of a few lines in our local weekly newspaper. I know Bill sure appreciated your kind note about how he cheered you up after your near crash. Old Bill showed your letter to everybody at the station and then some. He wasn't much for writing so had asked me to send you a verification and thanks. I admit he asked me three or four times and always I just said "Don't worry, I'll get to it." Bill passed on this past October and I hate to say I'm just getting around to sending this letter. Anyway, I don't think it was any coincidence that you heard him that night. I guess you could say that it really put a nice cap on his radio career in this town and re-started Bill's once famous smile. What the Good Lord can do with a humble 250 watts when he wants to reach somebody's heart!

Sincerely,
Jim Pierce, Program Director

A Smudged Postmark And No Return Address

If you've ever misplaced a letter, only to find it years later, you'll understand my joyful squeal when flipping open a long dormant *Broadcasting Yearbook* and happening upon something just right for this month's second small AM radio station epoch. Inside the plain business-size envelope clumsily stamped with what sort of looks like the "Z" of an Arizona postmark, was a single sheet article from the September 1997 *Guideposts* magazine. The piece, titled "*My Computer Prayer*" by Lois Lonnquist of Helena, Montana, had little to do with keyboards, monitors, and hard drives. Rather, it focused on a previously defunct standard broadcast facility that the author and her husband had just returned to the airwaves surrounding Great Falls, Montana.

Reportedly, during the winter of 1987, Mrs. Lonnquist's husband, Del, was growing restless after selling their two Midwestern radio outlets and attempting to be retired. Like many authentic local radio guys, he quickly missed what he knew best — hometown station ownership/operation. That led to perusing media brokers' prospectuses and eyeing that Class IV, single-kilowatt Great Falls station that had gone off the air and was down for the FCC count of *12 months dark and get deleted*. Mrs. Lonnquist wasn't convinced that this AM should figure into their plans, but, as she recalls, "We had 25 years of experience in radio. So, Del and I prayed for guidance, put our home up for sale, and made our move." Once in Great Falls, though, she wasn't particularly impressed with their new investment, writing, "We found a weathered one-story building surrounded by knee-deep prairie grass, and a chilly wind whistled through the silent station's 240-foot tower greatly in need of paint." As any radio man might imagine, husband Del didn't see things so gloomily. "We're going to love it here," he exclaimed to his skeptical spouse. Del Lonnquist then used some of that enthusiasm and a hammer to "pry-off some boards from a porch window [through which they] climbed for their first glimpse of the sad AM's dusty microphones and turntables." It took the couple nearly a month to rig-up the gear for re-introduction into the rest of their revitalization project carrying the call letters, KXGF.

Mailed to a hardware store on July 2, 1954, this business envelope from 560 KMON survives as one of the few pieces of early Eisenhower-era Great Falls radio memorabilia floating around on the Web. It is likely to have contained an invoice seeking payment for advertising on the 5-kilowatt outlet then owned by the Montana Farmer-Stockman *agricultural publication. The original KMON construction permit was issued in 1946 to Sun River Broadcasting Company for a 5,000-watt AM on 970 kilocycles, and construction of the station began early the next year. It quickly signed with ABC for network affiliation.*
Jan Lowry's research showed that before KMON 970 was ever completed, however, its "permit was sold by Sun River Broadcasting to the Montana Farmer Broadcasting Corporation. That firm then turned in its own permit for partially-built KMFR 560 in Great Falls. The KMON permit was modified by the FCC from 970 to the more favorable 560 kilocycle dial position ... with three, 440-foot-tall, guyed towers situated five miles south of the city and overlooking the Missouri River." Logically, a reason for this amalgamation of uncompleted CPs was to eliminate some potentially stiff competition in a relatively small media market.

The Lonnquist's — including a son, Roger, who served as engineer — decided that a primarily satellite-delivered big band/nostalgia music focus and local information would play well in Great Falls, so instituted that brand of mature programming. "Listeners loved the format," Lois noted, "but many advertisers were wary and waited for us to prove ourselves."

This lack of anticipated ad revenue got exacerbated by "payments from the

buyers of their Midwestern stations that were slow in coming." The couple essentially lived at KXGF 24/7 in order to run it parsimoniously. Several months into this seemingly unending marathon, she sat at her modest station's business office computer and typed a letter to God. The missive concluded with the urgent question, "How long can we go on?" Mostly, she wondered why they had felt directed to put KXGF back together. These queries got followed by a strong depression onto the question mark key that accented Lois' quandary with a monitor filled with: ??????????

The Fan Letter Made In Heaven

Lois Lonnquist's mood wasn't much brighter the next morning as she negotiated a stack of mail predominantly comprised of bills and too few checks from advertisers. The last envelope in that pile, however, contained a simple note from an elderly man who wanted the folks at KXGF to know that he and his wife were instant fans of the reborn station.

"We've been so happy to hear its old familiar tunes." The next paragraph told of his wife's recent illness and how KXGF kept them both in pleasant company, especially when his wife was awake in the wee hours battling pain. "Your music soothed her," he thanked the "dear friends at KXGF. And then early this morning, your announcer played *I'll Be Seeing You*, our favorite song. My wife looked at me and smiled," the listener recounted before closing with, "A short time later, she passed away."

Suddenly, Lois Lonnquist was filled with renewed appreciation for her radio station's ability to reach people and make a positive difference in their lives, even if only for a little while. She felt that the reminder had been an answer to a prayer. By the end of the year, a group approached the Lonnquist's with a respectable offer for KXGF, and the family accepted it. The *Guideposts* piece ends with Lois indicating that she and her husband later found a couple more stations to buy. Out of all of those radio projects, though, she smiles, "KXGF is our favorite."

A Hobson's Choice Of Hertzian Waves

For at least a couple of decades, even the most fickle Great Falls area radio listener would have to admit that the process of picking his or her best-loved local station was a no brainer. That's because — with the fleeting early '20s exception of KDYS, a 50-watt experiment of the *Great Falls Tribune* newspaper — from 1922 until after World War II, radio fans in the Montana city of about 50,000 had but one hometown audio voice, KFBB (now KEIN 1310 5-kilowatts day/1-kilowatt night). Great Falls broadcast buffs anticipated a treat, however, when learning of the extreme aural expansion planned for their community circa 1947-48.

During that brief period, the *Broadcasting Yearbook* not only noted KFBB, but four full time AM construction permits, each with all of the serious promise of call letters and imminent on-air target dates: KMFR (560 kilocycles, 5 kilowatts), KMON (790 kilocycles, 5 kilowatts), KGFM (1150 kilocycles, 5 kilowatts), and KXLK (1400 kilocycles, 250 watts).

As things shook down and the prospective station owners reconsidered their likely return on investment, two of the 5-kilowatt permit holders merged — resulting in KMFR on 560 kilohertz — and the 790 CP never actually materializing. Nor did KGFM go beyond paper. At just a quarter kilowatt and a 150-foot rooftop tower, the smallest of the post-war lot, KXLK, beat these others to be KFBB's first real competitor. Officially, KXLK was FCC-autho-

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Had this linen postcard of the Park Hotel been published about 10 years after its supposed late 1930s rendering, the picture would have included signage denoting tenant KXLK, predecessor to KXGF, the modest AM survivor in our story. No doubt, too, there's be pictorial evidence of a bit of the Great Falls station's tower pointing north from the five story venue's roof.

rized in late June 1946 as KSTR. But by its March 1947 debut, the little AM was called KXLK to reflect the "XL" family of sister Montana radio outlets, KXLF Butte, KXLJ Helena, KXLL Missoula, KXLO Lewiston, and KXLQ Bozeman.

The Park Hotel at the corner of Great Fall's Central Avenue and Park Drive served as KXLK's first home. Jan Lowry's research details that the establishment not only housed the station's studio and business offices, but also contained its transmitter (in the basement) and antenna on the roof. Not much out of the ordinary shows up on KXLK's history until a 1958 sale and callsign change in 1960 to KARR. Nearly coinciding with this new identity were an early 1961 daytime power increase to a kilowatt, as well as a new Great Falls studio/transmitter venue at 1900 Third Street, N.W. Jan cites an "upbeat Middle-of-the-Road music format on KARR from 5:30 a.m. until 2 a.m. daily" circa 1971, plus a litany of ownership transfers in 1973, 1977, 1982, 1986 (when soon thereafter, it apparently went dark), 1987, 1989, 1995, and 2006. Jan also notes a 1989 studio shift to 1300 Central Avenue, West. Along the way, KARR became KYOT (in summer 1984), and then KXGF during 1987. In addition to the longtime M-O-R offering, 1400 kilohertz in Great Falls once featured a religious format, "Hit Album Rock Music," C&W tunes, big bands/standards, hot adult contemporary, and sports talk programming KXGF uses its 680 watts (mandated in a 2000 tower move to a new Fox Road site just south of Great Falls) in order to broadcast today.

That Small Station In The Little *Guideposts* Magazine

Though not mentioned specifically in Jan's FCC-mined data, it appears that the Lonnquist's contribution to the resilient 1400 kilohertz Montana survivor was their 1987 re-casting it as KXGF after believing the station to have worth beyond its modest power. While they didn't keep the station very long, the radio family succeeded in resurrecting KXGF's signal so that it could bring a bit of joy to whomever needed the friendship available within the reach of its fleeting radio waves.

And so ends another broadcast day from the Radio Drama studios.

unwired

BY RICHARD FISHER, KPC6PC

The Weirder Side of Wireless, and Beyond

Au Naturel: 'No Antenna? Noisy Location? No Problem!

OK, imagine you're in the most remote, radio-quiet spot on the planet and want to hear what the rest of us city slickers can't. Put on your *"I'm Gloating"* hat and link to <http://www.ab9il.net/vlf>, "Natural Radio VLF Listening Via the Internet" where you can bask in some of the best RF that's ever bombarded Earth.

*(**HOUSEHOLD HINT:** Why invest in a "white noise" generator when you've got this? Crank it up and go zzzzzzzzzzzzzzzzzz ... – KPC6PC.)*

FCC: 'We Can Hear You Now'

What price ambience? It could ring up to more than $100K for Perfect Cuts Salon in San Antonio, Texas.

It seems the Federal Communications Commission does not like the cut of Ronald Bethany's lights, which it alleges have been putting out a signal on 705 MHz. Did we mention that's the frequency used for AT&T's 4G services? Did we mention its cell site *is right next door?*

AT&T tipped off the FCC which not-so-subtly warned it might take a little off the top of Perfect Cuts.

"If, after receipt of this citation Mr. Bethany again violates Section 15.5(b) of the Rules by engaging in conduct of the type described herein, the Commission may impose monetary forfeitures not to exceed $16,000 for each such violation or each day of a continuing violation, and up to $112,500 for any single act or failure to act."

There hasn't been a battle involving hair this fierce since the Beatles invasion of 1964.

*(**READ:** The FCC citation at <http://bit.ly/1cxRlbP>. – KPC6PC.)*

Drawing a Blanc: 'You Bend It, We Mend It'

Bugs Bunny, Daffy Duck, Porky Pig, Tweety Bird, Sylvester the Cat, Yosemite Sam, Foghorn Leghorn, Marvin the Martian, Pepé Le Pew, Speedy Gonzales, Wile E. Coyote, the Tasmanian Devil, and lots of other characters from Looney Tunes and Merrie Melodies popped up on Los Angeles-area Citizens Band when Mel Blanc was on the air.

But that wasn't the only radio "show" in which he starred. For a year beginning in 1946, "The Mel Blanc Show" featured "the man of 1,000 voices" as the hapless owner of "Mel's Fix-It Shop."

Listen to 42 knee-slapping episodes free online at <http://bit.ly/16YKwgW>.

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By Juergen A. Weigl, OE5CWL

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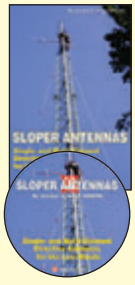

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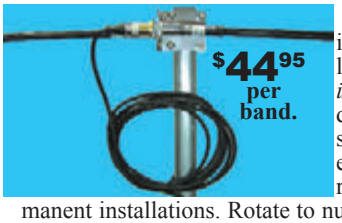

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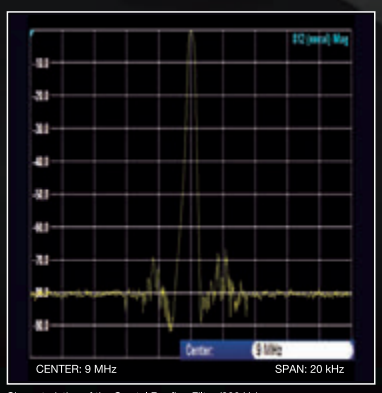

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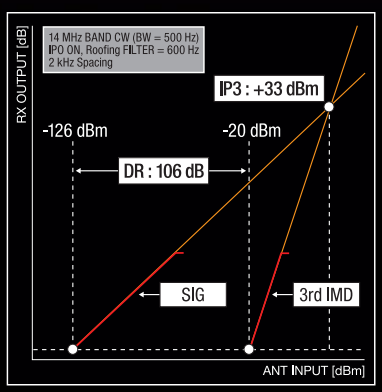

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